

MMSE-Based Power Control for Amplify-and-Forward Relay Systems in Time-Varying Channels

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Abstract

This study proposes a novel power control method for amplify-and-forward (AF) relay systems operating in time-varying channels. Transmit power control between the source and relay nodes significantly enhances the performance of the AF relay system, and the improvements are proportional to the number of antennas. However, these enhancements are restricted by the presence of time-varying channels, and this limitation increases in severity with increasing number of antennas. To address the challenges posed by these channel variations, the proposed method adaptively optimizes the power-scaling factors to minimize the mean-squared errors under a power-inequality constraint. Numerical results demonstrate that the proposed method effectively mitigates the bit-error-rate performance degradation caused by channel variations, maintaining robust performance even in systems with a large number of antennas. This approach offers a promising solution for improving the reliability and efficiency of AF relay systems under dynamic channel conditions.

Index Terms: Amplify-and-forward relay, Minimum mean-squared error (MMSE), Power control, Time-varying channels

I. INTRODUCTION

The number of mobile devices has been increasing over the last few decades, and this trend will continue due to the proliferation of mobile broadband services. Consequently, mobile networks now require higher data rates and wider coverage areas than were previously required [1]. Traditionally, relays have effectively enhanced the data rate and extended the coverage of mobile networks without requiring an additional wired backhaul [2]. Accordingly, mobile network standards such as IEEE 802.16, long-term evolution, and new radio have adopted relay systems.

Relaying between a source node (SN) and destination node (DN) can be accomplished via several schemes, which are distinguished by the signal processing of a relay node (RN). The most straightforward relaying scheme is amplify-and-

forward (AF), in which the signals received at the RN are linearly processed and retransmitted to the DN [3]. Owing to the absence of a decoding procedure at the RN, the AF scheme is less complex than other strategies such as the decode-and-forward and compressed-and-forward schemes. However, the noise amplification induced by the RN limits the AF scheme and causes performance degradation. To circumvent this problem, various power control methods based on channel information in AF relay systems have been adopted [3]. For example, a power control framework was developed to maximize the signal-to-noise ratio (SNR) of the DN [4]. In [5], the optimal power-scaling factors of the RN were represented using two objective functions: the minimum mean-squared error (MSE) and zero forcing. Further, power control methods based on the minimum MSE (MMSE) criterion under power-inequality constraints were designed

Received 12 June 2024, Revised 13 July 2024, Accepted 16 July 2024

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Open Access <https://doi.org/10.56977/jicce.2024.22.4.267>

print ISSN: 2234-8255 online ISSN: 2234-8883

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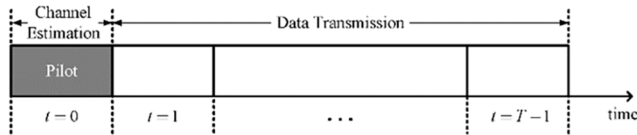


Fig. 1. Frame structure with one pilot and $T-1$ data symbols.

[6]. Channel information is uncertain in AF relay systems because it is obtained from channel estimation at an SN or DN (and feedback from them). Thus, channel uncertainty should be carefully considered when designing power controls, particularly when the channel varies over time. However, all the previously mentioned studies assumed time-invariant channels throughout the data transmission. In time-varying channel environments, the power control performance degrades owing to outdated information. Therefore, channel variations should be carefully considered to avoid significant performance degradation.

In this study, we propose a power control method for AF relay systems in time-varying channels. Closed-form expressions of the optimal power-scaling factors are derived from the MMSE problem under a local power-inequality constraint. However, obtaining a solution is challenging because the optimal power-scaling factors are heavily entangled. To obtain a feasible solution, we employ an alternating-optimization-based iterative algorithm that sequentially optimizes the power-scaling factors while keeping all others fixed. The simulation results confirm that the proposed power control method effectively mitigates bit-error-rate (BER) performance degradation caused by channel variations when each node has four or more antennas.

The remainder of this paper is organized as follows: The system and signal models for the considered AF relay system and time-varying channels are introduced in Section II, while the proposed power control method is presented in Section III. The simulation results and conclusions are presented in Section IV and Section V, respectively. The following notation is used throughout the paper: boldface lower-case and capital letters denote vectors and matrices, respectively, while lower-case letters denote scalars; superscript H represents a conjugate transpose; $|x|$ and $\|x\|$ denote the absolute value of x and the 2-norm of x , respectively; I_n represents an $n \times n$ identity matrix; $\mathbb{E}\{\cdot\}$ is the expectation of a random variable or matrix; $\mathcal{CN}(0, \sigma^2)$ represents a zero-mean complex Gaussian distribution with variance σ^2 ; $(x)^+$ denotes $(x + |x|)/2$ for an arbitrary real number x ; and $\Re(x)$ is the real part of x .

II. SYSTEM AND SIGNAL MODELS

In this study, we consider downlink communication in a

two-hop half-duplex AF relay system with scaled eigen beamforming [6]. The SN, RN, and DN are equipped with N_S , N_R , and N_D antennas, respectively. For beamforming purposes, N_S and N_D are greater than or equal to two, while N_R has no constraints. The direct path between the SN and DN is not considered. Denoting T as the number of transmitted symbols per frame, Fig. 1 illustrates the frame structure, which consists of one pilot and $T-1$ data symbols. The pilot symbol is transmitted at the first transmission time. The channels are estimated separately at the RN and DN and then fed back to every node through a broadcasting channel. In this study, perfect channel feedback is assumed. After the channel estimation and feedback completed, the remaining transmission time is used to transmit $T-1$ data symbols.

Let $\mathbf{H}(t) \in \mathbb{C}^{N_R \times N_S}$ and $\mathbf{F}(t) \in \mathbb{C}^{N_D \times N_R}$ denote the first- and second-hop channel matrices at the t th symbol time, respectively, for symbol time index $t \in \{0, \dots, T-1\}$. The channel model conforms to Rayleigh fading, where the elements of $\mathbf{H}(t)$ and $\mathbf{F}(t)$ are independent random variables following $\mathcal{CN}(0, 1)$. The first-order Gauss-Markov process is assumed to model the time variation of the first- and second-hop channels as follows [7]:

$$\mathbf{H}(t) = \rho_h \mathbf{H}(t-1) + \mathbf{Z}_h(t) = \rho_h^t \mathbf{H}(0) + \sum_{i=1}^{t-1} \rho_h^{t-i} \mathbf{Z}_h(i), \quad (1)$$

$$\mathbf{F}(t) = \rho_f \mathbf{F}(t-1) + \mathbf{Z}_f(t) = \rho_f^t \mathbf{F}(0) + \sum_{i=1}^{t-1} \rho_f^{t-i} \mathbf{Z}_f(i). \quad (2)$$

Here, the elements of $\mathbf{Z}_h(t) \in \mathbb{C}^{N_R \times N_S}$ and $\mathbf{Z}_f(t) \in \mathbb{C}^{N_D \times N_R}$ are independent random variables following $\mathcal{CN}(0, 1-\rho_h^2)$ and $\mathcal{CN}(0, 1-\rho_f^2)$, respectively, where $\rho_h \in [0, 1]$ and $\rho_f \in [0, 1]$ are the first- and the second-hop correlation coefficients, respectively. In this study, we assume that ρ_h and ρ_f are known at each node because the channel correlation coefficient can be accurately estimated [7].

For the beamforming, all nodes should know the perfect channel information (i.e., $\mathbf{H}(t)$ and $\mathbf{F}(t)$ for $t \in 0, \dots, T-1$). However, only the estimates of $\mathbf{H}(0)$ and $\mathbf{F}(0)$, denoted by $\hat{\mathbf{H}}(0) \in \mathbb{C}^{N_R \times N_S}$ and $\hat{\mathbf{F}}(0) \in \mathbb{C}^{N_D \times N_R}$, respectively, are available. The estimated channel matrices are modeled as:

$$\hat{\mathbf{H}}(0) = \mathbf{H}(0) + \Delta_h, \quad (3)$$

$$\hat{\mathbf{F}}(0) = \mathbf{F}(0) + \Delta_f, \quad (4)$$

where the elements of $\Delta_h \in \mathbb{C}^{N_R \times N_S}$ and $\Delta_f \in \mathbb{C}^{N_D \times N_R}$ are independent random variables following $\mathcal{CN}(0, \sigma_h^2)$ and $\mathcal{CN}(0, \sigma_f^2)$, respectively. Consequently, the relay processing matrix and source-destination beamforming vectors are obtained from $\hat{\mathbf{H}}(0)$ and $\hat{\mathbf{F}}(0)$, as follows [6]:

$$\mathbf{W}(t) = \alpha(t) \mathbf{v}_f \mathbf{u}_h^H \in \mathbb{C}^{N_R \times N_R}, \quad (5)$$

$$\mathbf{a}(t) = \beta(t) \mathbf{v}_h \in \mathbb{C}^{N_S \times 1}, \quad (6)$$

$$\mathbf{b}(t) = \gamma(t) \mathbf{u}_f \in \mathbb{C}^{N_D \times 1}. \quad (7)$$

Here, $\mathbf{u}_h \in \mathbb{C}^{N_R \times 1}$ and $\mathbf{v}_h \in \mathbb{C}^{N_S \times 1}$ are the left and right singular vectors, respectively, corresponding to the largest sin-

gular value η_h of $\hat{\mathbf{H}}(0)$; $\mathbf{u}_f \in \mathbb{C}^{N_D \times 1}$ and $\mathbf{v}_f \in \mathbb{C}^{N_R \times 1}$ are the left and right singular vectors, respectively, corresponding to the largest singular value η_f of $\hat{\mathbf{F}}(0)$. The power-scaling factors $\alpha(t) \in \mathbb{R}$, $\beta(t) \in \mathbb{R}$, and $\gamma(t) \in \mathbb{R}$ are designed in Section III.

The SN transmits the data over the first-hop link. The t th transmitted symbol vector of the SN is expressed as

$$\mathbf{x}_s(t) = \mathbf{a}(t)d(t) \in \mathbb{C}^{N_S \times 1}, \quad (8)$$

where the t th data symbol is denoted by $d(t) \in \mathbb{C}$ with $\mathbb{E}\{|d(t)|^2\} = 1$. At the RN, the t th received symbol vector is given by

$$\mathbf{y}_R(t) = \mathbf{H}(t)\mathbf{x}_s(t) + \mathbf{n}_R(t) \in \mathbb{C}^{N_R \times 1}. \quad (9)$$

Here, $\mathbf{n}_R(t) \in \mathbb{C}^{N_R \times 1}$ is an additive white Gaussian noise (AWGN) vector with $\mathbb{E}\{\mathbf{n}_R(t)\mathbf{n}_R^H(t)\} = \sigma_R^2 \mathbf{I}_{N_R}$. The RN performs linear processing of $\mathbf{y}_R(t)$ with $\mathbf{W}(t)$ as follows:

$$\mathbf{x}_R(t) = \mathbf{W}(t)\mathbf{y}_R(t) \in \mathbb{C}^{N_R \times 1}. \quad (10)$$

Then, $\mathbf{x}_R(t)$ is transmitted to the DN over the second-hop link. The t th received symbol vector at the DN is given by

$$\mathbf{y}_D(t) = \mathbf{F}(t)\mathbf{x}_R(t) + \mathbf{n}_D(t) \in \mathbb{C}^{N_D \times 1}, \quad (11)$$

where $\mathbf{n}_D(t) \in \mathbb{C}^{N_D \times 1}$ is an AWGN vector with $\mathbb{E}\{\mathbf{n}_D(t)\mathbf{n}_D^H(t)\} = \sigma_D^2 \mathbf{I}_{N_D}$. Substituting (8)-(10) into (11), the DN combines $\mathbf{y}_D(t)$ to obtain the t th estimated data symbol $\hat{d}(t) = \mathbf{b}^H(t)\mathbf{y}_D(t) \in \mathbb{C}$, which is expressed as

$$\hat{d}(t) = \mathbf{b}^H(t)\mathbf{F}(t)\mathbf{W}(t)\mathbf{H}(t)\mathbf{a}(t)d(t) + \mathbf{b}^H(t)\mathbf{F}(t)\mathbf{W}(t)\mathbf{n}_R(t) + \mathbf{b}^H(t)\mathbf{n}_D(t). \quad (12)$$

III. POWER CONTROL METHOD IN TIME-VARYING CHANNELS

In this section, we design the power-scaling factors $\{\alpha(t), \beta(t), \gamma(t)\}$ to minimize the MSE between $d(t)$ and $\hat{d}(t)$ under the local inequality constraint. Henceforth, the time index t is omitted whenever convenient.

The transmit power of the SN and RN are independently bounded by P_S and P_R (i.e., $\mathbb{E}\{|\mathbf{x}_S|^2\} \leq P_S$ and $\mathbb{E}\{|\mathbf{x}_R|^2\} \leq P_R$), respectively, under the local power-inequality constraint. The MMSE of the data symbol is expressed as follows:

$$\begin{aligned} \min_{\alpha_L, \beta_L, \gamma_L} \quad & \mathbb{E}\{|d - \hat{d}|^2\}, \\ \text{s.t.} \quad & \mathbb{E}\{|\mathbf{x}_S|^2\} \leq P_S \text{ and } \mathbb{E}\{|\mathbf{x}_R|^2\} \leq P_R \end{aligned} \quad (13)$$

where α_L , β_L , and γ_L are the power-scaling factors under the local power-inequality constraint. Using the Lagrange multiplier method and substituting (8), (10), and (12) into (13), we obtain

$$\min_{\alpha_L, \beta_L, \gamma_L, \lambda_S, \lambda_R} J_L, \quad (14)$$

where λ_S and λ_R are the non-negative Lagrangian multipliers, and J_L is the cost function under the local power-inequality constraint, which is given by

$$\begin{aligned} J_L = & \mathbb{E}\{|d - \hat{d}|^2\} + \lambda_S(\mathbb{E}\{|\mathbf{x}_S|^2\} - P_S) + \lambda_R(\mathbb{E}\{|\mathbf{x}_R|^2\} - P_R) \\ = & 1 - 2\alpha_L\beta_L\gamma_L\rho_h^t\eta_h\rho_f^t\eta_f + \alpha_L^2\beta_L^2\gamma_L^2\rho_h^{2t}\eta_h^{2t}\rho_f^{2t}\eta_f^{2t} \\ & + \alpha_L^2\gamma_L^2\rho_f^{2t}\eta_f^{2t}\sigma_R^2 + \gamma_L^2\sigma_D^2 \\ & + \lambda_S(\beta_L^2 - P_S) + \lambda_R(\alpha_L^2\beta_L^2\rho_h^{2t}\eta_h^{2t} + \alpha_L^2\sigma_R^2 - P_R). \end{aligned} \quad (15)$$

Here, the corrupted singular values $\tilde{\eta}_h$ and $\tilde{\eta}_f$ satisfy

$$\begin{aligned} \tilde{\eta}_h^2 = & \mathbb{E}\{|\eta_h - \mathbf{u}_h^H \Delta_h \mathbf{v}_h + \sum_{i=1}^t \rho_h^{-i} \mathbf{u}_h^H \mathbf{Z}_h(i) \mathbf{v}_h|^2\} \\ = & \eta_h^2 + 2\eta_h \mathbb{E}\{\Re(\delta_h)\} + \mathbb{E}\{|\delta_h|^2\} + \rho_h^{-2t} - 1 \in \mathbb{R}, \end{aligned} \quad (16)$$

$$\begin{aligned} \tilde{\eta}_f^2 = & \mathbb{E}\{|\eta_f - \mathbf{u}_f^H \Delta_f \mathbf{v}_f + \sum_{i=1}^t \rho_f^{-i} \mathbf{u}_f^H \mathbf{Z}_f(i) \mathbf{v}_f|^2\} \\ = & \eta_f^2 + 2\eta_f \mathbb{E}\{\Re(\delta_f)\} + \mathbb{E}\{|\delta_f|^2\} + \rho_f^{-2t} - 1 \in \mathbb{R}, \end{aligned} \quad (17)$$

where δ_h and δ_f are defined as $-\mathbf{u}_h^H \Delta_h \mathbf{v}_h$ and $-\mathbf{u}_f^H \Delta_f \mathbf{v}_f$, respectively. By equating the derivatives of J_L with respect to the optimization variables to zero, the optimal power-scaling factors and Lagrangian multipliers under the local power-inequality constraint are expressed as

$$\alpha_L = \frac{\beta_L \gamma_L \rho_h^t \eta_h \rho_f^t \eta_f}{(\gamma_L^2 \rho_L^2 \tilde{\eta}_f^2 + \lambda_R)(\beta_L^2 \rho_h^2 \tilde{\eta}_h^2 + \sigma_R^2)}, \quad (18)$$

$$\beta_L = \frac{\alpha_L \gamma_L \rho_h^t \eta_h \rho_f^t \eta_f}{\alpha_L^2 \gamma_L^2 \rho_h^{2t} \tilde{\eta}_h^{2t} \rho_f^{2t} \tilde{\eta}_f^{2t} + \lambda_R \alpha_L^2 \rho_h^{2t} \tilde{\eta}_h^{2t} + \lambda_S}, \quad (19)$$

$$\gamma_L = \frac{\alpha_L \beta_L \rho_h^t \eta_h \rho_f^t \eta_f}{\alpha_L^2 \beta_L^2 \rho_h^{2t} \tilde{\eta}_h^{2t} \rho_f^{2t} \tilde{\eta}_f^{2t} + \alpha_L^2 \sigma_R^2 \rho_f^{2t} \tilde{\eta}_f^{2t} + \sigma_D^2}, \quad (20)$$

$$\lambda_R = \left(\frac{\beta_L \gamma_L \rho_h^t \eta_h \rho_f^t \eta_f}{\sqrt{P_R(\beta_L^2 \rho_h^{2t} \tilde{\eta}_h^2 + \sigma_R^2)}} - \gamma_L \rho_f^{2t} \tilde{\eta}_f^{2t} \right)^+, \quad (21)$$

$$\lambda_S = \left(\frac{\alpha_L \gamma_L \rho_h^t \eta_h \rho_f^t \eta_f}{\sqrt{P_S}} - \alpha_L \rho_h^{2t} \tilde{\eta}_h^{2t} (\gamma_L^2 \rho_f^{2t} \tilde{\eta}_f^{2t} + \lambda_R) \right)^+. \quad (22)$$

Direct computation of the optimal solution for (18)-(22) is challenging because $\alpha_L(t)$, $\beta_L(t)$, $\gamma_L(t)$, $\lambda_S(t)$, and $\lambda_R(t)$ are functions of other optimization variables. Therefore, we adopt the alternating optimization presented in Algorithm 1. Here, the subscript k denotes the iteration index. The maximum number of iterations and threshold for the stopping criterion are denoted by $K_{\max}$ and ϵ , respectively. In line 7, $\{\alpha_{L,k}(t), \beta_{L,k}(t), \gamma_{L,k}(t)\}$ is updated by inputting the power-

scaling factors and Lagrangian multipliers obtained at the $(k-1)$ th iteration into (18)-(20). When implementing Algorithm 1, an infeasible solution $\alpha_{L,k}(t) = \beta_{L,k}(t) = \gamma_{L,k} = \infty$ is obtained if $\beta_{L,0}(t)$ and $\gamma_{L,0}(t)$ are initialized by zero. To resolve this issue, $\{\beta_{L,0}(t), \gamma_{L,0}(t), \lambda_{S,0}(t), \lambda_{R,0}(t)\}$ are set to $\{1, 1, 0, 0\}$ at $t = 1$.

Algorithm 1: Power control algorithm under local power-inequality constraint.

```

1 Input: Correlation coefficients  $\{\rho_h, \rho_f\}$ , estimated
   singular values  $\{\eta_h, \eta_f\}$ , corrupted singular values
    $\{\tilde{\eta}_h(t), \tilde{\eta}_f(t)\}$ , and local power constraints  $\{P_S, P_R\}$ .
2 Output: Relay processing matrix and source-
   destination beamforming vectors  $\{W(t), a(t), b(t)\}$ .
3 for  $t = 1$  to  $T - 1$  do
4    $\beta_{L,0}(t) = \gamma_{L,0}(t) = 1$  and  $\lambda_{S,0}(t) = \lambda_{R,0}(t) = 0$ .
5   Initialize  $J_{L,0}(t) = 0$ .
6   for  $k = 1$  to  $K_{\max}$  do
7     Update  $\alpha_{L,k}(t)$ ,  $\beta_{L,k}(t)$ , and  $\gamma_{L,k}(t)$  as (18)–(20).
8     Obtain  $\lambda_{R,k}(t)$  and  $\lambda_{S,k}(t)$  from (21) and (22).
9     Compute  $J_{L,k}(t)$  from (15).
10    if  $0 \leq J_{L,k-1}(t) - J_{L,k}(t) \leq \epsilon$  then
11      Break.
12    end if
13  end for
14   $W(t) = \alpha_{L,k}(t)v_f u_h^H$ ,  $a(t) = \beta_{L,k}(t)v_h$ ,
      $b(t) = \gamma_{L,k}(t)u_f$ .
15 end for

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IV. SIMULATION RESULTS

The performance of the proposed power control algorithm is evaluated through numerical simulations. The simulation parameters are listed in Table 1. The local power-inequality constraint is set to $P_S = P_R = 1$. For the simulation, the received SNRs at the RN and DN are defined as P_S/σ_R^2 and P_R/σ_D^2 , respectively [6]. Because the radio link between the SN and RN is usually in good condition during downlink communication [8], P_S/σ_R^2 is fixed at 25 dB. The data symbols are modulated using 16-quadrature amplitude modulation (QAM). The channel correlation coefficient typically ranges between 0.9 and 0.99 [9]. Accordingly, the correlation coefficients ρ_h and ρ_f are set to 0.99 and 0.97, respectively. The maximum number of iterations $K_{\max}$ and threshold value ϵ are 1000 and 10^{-4} , respectively. The variance of channel estimation errors σ_h^2 and σ_f^2 is fixed at 0.1. From (18)–(22), the square of the corrupted singular values $\tilde{\eta}_h^2(t)$ and $\tilde{\eta}_f^2(t)$ are required to obtain the power-scaling factors. Thus, assuming channel uncertainty of up to 10% (i.e., $\sigma_h^2 \leq 0.1$ and $\sigma_f^2 \leq 0.1$), the following model is employed [6]:

Table 1. Simulation parameters for verifying the performance of the proposed algorithm

Parameters	Values
Local power-inequality constraint	$P_S = P_R = 1$
SNR at RN	$P_S/\sigma_R^2 = 25$ dB
Data symbol modulation	16-QAM
Channel correlation coefficients	$\{\rho_h, \rho_f\} = \{0.99, 0.97\}$
Maximum number of iterations	$K_{\max} = 1000$
Threshold for stopping criterion	$\epsilon = 10^{-4}$
Variance of channel estimation error	$\sigma_h^2 = \sigma_f^2 = 0.1$

Table 2. Numerical values of $\{a_2, a_1, b_2, b_1\}$ in (23)

$\{N_S, N_R, N_D\}$	a_2	a_1	b_2	b_1
$\{2, 2, 2\}$	0.7961	-1.8052	2.1372	1.0164
$\{4, 4, 4\}$	1.3696	-3.0867	7.4859	1.0584
$\{8, 8, 8\}$	2.1750	-4.8452	19.3744	1.1544

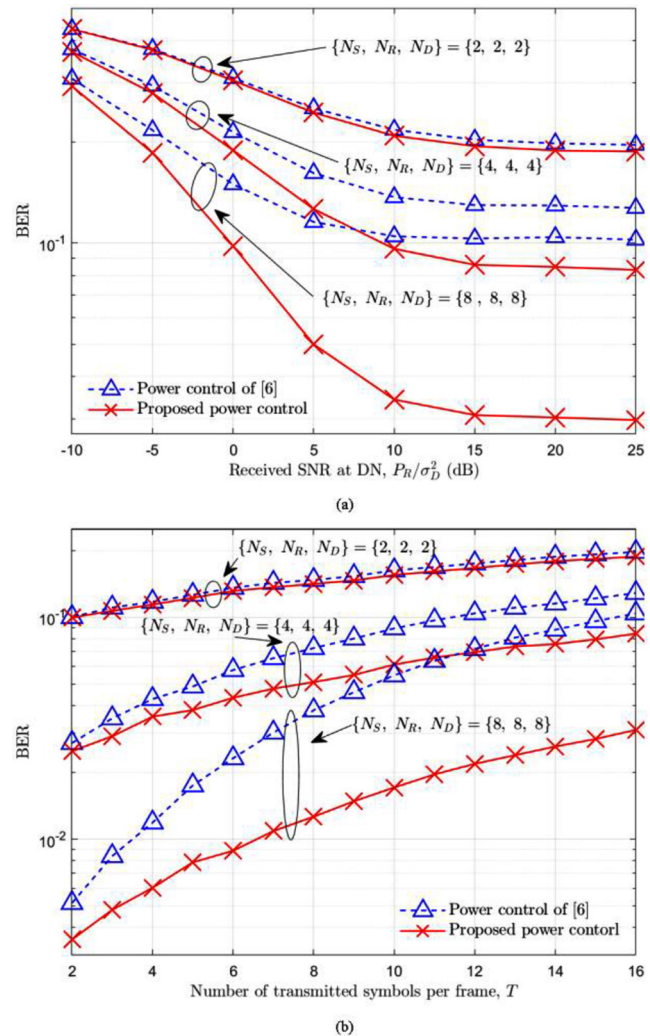


Fig. 2. 16-QAM BER performance comparison with $P_S/\sigma_R^2 = 25$ dB. (a) BER over P_R/σ_D^2 for $T=16$. (b) BER over T for $P_R/\sigma_D^2 = 20$ dB.

$$\mathbb{E}\{\Re(\delta)\} \approx (a_2\sigma^2 + a_1)\sigma^2, \quad (23a)$$

$$\mathbb{E}\{|\delta|^2\} \approx (b_2\sigma^2 + b_1)\sigma^2, \quad (23b)$$

where the coefficients $\{a_2, a_1, b_2, b_1\}$ are numerically computed using the polynomial fitting method. Table 2 lists the values of $\{a_2, a_1, b_2, b_1\}$ used in the simulations. By substituting (23) into (16) and (17), $\tilde{\eta}_h^2(t)$ and $\tilde{\eta}_f^2(t)$ are obtained.

In Fig. 2, the BER performance is evaluated over P_R/σ_D^2 and T to demonstrate the robustness of the proposed power control method against time-varying channels. For comparison, the power control method in [6] is adopted as a baseline method, in which the power-scaling factors obtained at $t = 0$ are applied for $t \geq 1$. The simulation results show that the performance gap between the proposed and baseline methods is negligible when the number of antennas at each node is two. In contrast, the proposed method effectively mitigates BER performance degradation caused by the time-varying channels when the number of antennas at each node is four or eight. Particularly, the proposed method with $\{N_S, N_R, N_D\} = \{4, 4, 4\}$ outperforms the baseline method with $\{N_S, N_R, N_D\} = \{8, 8, 8\}$ when $P_R/\sigma_D^2 \geq 10$ dB or $T \geq 12$. The baseline method exhibits limited BER performance because the power-scaling factors are not adjusted to the channel variations. In contrast, the proposed method adaptively controls the power-scaling factors at each symbol time to minimize the MMSE of the data symbols at the DN.

V. CONCLUSION

In this study, we proposed a power control method under the local power-inequality constraint for reliable communication in multi-input multi-output amplify-and-forward relay systems in time-varying channels. In the proposed method, the power-scaling factors of each node were obtained from an alternating optimization-based iterative algorithm. The simulation results demonstrated that the proposed method effectively mitigated the BER performance degradation caused by time-varying channels when each node had more than four antennas.

ACKNOWLEDGEMENTS

This work was supported in part by the Institute of Information and Communications Technology Planning and Evaluation (IITP) Grant funded by the Korea Government

(MSIT) (No.2021-0-00874, Development of Next Generation Wireless Access Technology Based on Space Time Line Code, 30%; No. 2022-0-00635, Development of 5G Industrial Terminal Technology Supporting 28 GHz Band/Private 5G Band/NR-U Band, 25%); in part by the National Research Foundation of Korea (NRF) Grant funded by the Korea Government (MSIT) under Grant 2022R1A2C1003 750 & RS-2024-00405510; and in part by the Chung-Ang University research grant in 2024.

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