Performance Analysis of a Dual-Hop Cooperative Relay Network with Co-Channel Interference

Min LIN [1, 2], *Heng WEI* [3], *Jian OUYANG* [4], *Kang AN* [3], *Can YUAN* [3]

[1] PLA University of Science and Technology, Nanjing, China
[2] National Mobile Communications Research Laboratory, Southeast University, Nanjing, China
[3] College of Communication Engineering, PLA University of Science and Technology, Nanjing, China
[4] Inst. of Signal Processing and Transmission, Nanjing University of Posts and Telecommunications, Nanjing, China

linmin63@163.com, weiheng63@163.com, ouyangjian@njupt.edu.cn, ankang@nuaa.edu.cn, yuancanchina@gmail.com

Abstract. *This paper analyzes the performance of a dual-hop amplify-and-forward (AF) cooperative relay network in the presence of direct link between the source and destination and multiple co-channel interferences (CCIs) at the relay. Specifically, we derive the new analytical expressions for the moment generating function (MGF) of the output signal-to-interference-plus-noise ratio (SINR) and the average symbol error rate (ASER) of the relay network. Computer simulations are given to confirm the validity of the analytical results and show the effects of direct link and interference on the considered AF relay network.*

Keywords

Amplify-and-forward, cooperative relay network, co-channel interference, performance analysis.

1. Introduction

Recently, the application of relay technology has attracted many researchers as a low-cost strategy to increase the capacity, reliability and coverage of wireless networks without requiring extra power or bandwidth. (See [1], [2] and the citations therein for example.) Among the commonly used relay protocols, such as amplify-and-forward (AF), decode-and-forward (DF) and compress-and-forward (CF), AF scheme is of particular interest due to its simplicity. When only one relay is exploited to amplify the signal, the configuration is called as dual-hop relay network.

When the channel state information (CSI) is available, the performance of dual-hop AF relay networks has been well studied in a large body of open literatures [3]-[5]. In practice, however, co-channel interference (CCI) due to frequency reuse often severely degrades the performance of wireless systems. Motivated by this fact, the authors of [6] have investigated the effect of CCI on a variable gain AF relay network over Rayleigh fading channels. Moreover, the authors in [7] have analyzed the ergodic capacity of AF relay system with CCI present at both relay and destination over Nakagami-*m* fading channels. By considering the joint effects of CCI and imperfect CSI, the performance of a dual hop variable gain AF relaying with BF has been studied in [8], where the outage probability (OP), average symbol error rate (ASER) and asymptotic analysis at high SNR have been presented.

An extension of the previous works to the multi-antenna AF relaying has been presented in [9]-[11], respectively. In [9], the authors have analyzed the performance of multi-antenna AF relay systems with feedback delay at the source and CCI at the relay. By assuming that the statistical CSI is available at the relay, the analytical expressions for the OP, Ergodic capacity, ASER of a multi-antenna AF relay network have been obtained in [10]. Moreover, both the closed-form and asymptotic expressions of OP for multi-antenna AF relaying with have derived in [11], where the more general case with both the relay and destination corrupted by multiple co-channel interferers has been considered. However, in the aforementioned studies, the direct link between the source and destination is assumed to be unavailable due to heavy shading, which is not an actual model in many practical environments.

In this paper, unlike the previous related works, we conduct the performance of a dual-hop AF cooperative relay network, where direct link exists between the source and destination, and multiple CCIs corrupt the received signals at the relay. In particular, we derive the new theoretical expressions for the moment generating function (MGF) of the output signal-to-interference-plus-noise ratio (SINR) as well as the average symbol error rate (ASER) of the relay system. To the best of our knowledge, this is the first time such analytical results have been presented.

2. System Model

As shown in Fig. 1, we consider a dual-hop AF relay network, where a source *S* communicates with a destination *D* through a relay *R*. We assume that each node is equipped with a single antenna, and there exists a direct link between *S* and *D*. Here, all the wireless channels are subject to Rayleigh fading. The complete communication takes place in two time slots. In the first time slot, *S* broadcasts the signal to *D* and *R* through the fading channels h_0 and h_1, respectively. Meanwhile, *R* is also impaired

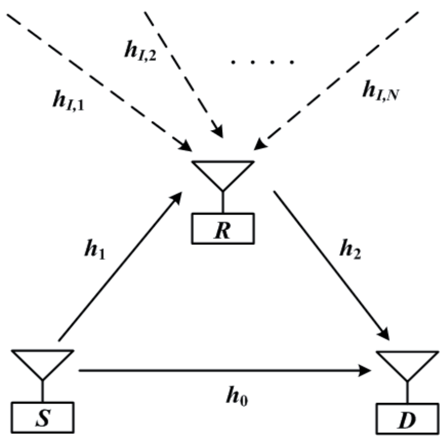

Fig. 1. System model of a dual-hop cooperative relay network with CCI.

by N CCIs through fading channel $h_{I,n}$ ($n = 1,2,\ldots,N$). As such, the received signals at D and R can be, respectively, expressed as

$$y_0(t) = \sqrt{P_S}\, h_0 x_S(t) + n_0(t), \tag{1}$$

$$y_1(t) = \sqrt{P_S}\, h_1 x_S(t) + \sum_{n=1}^{N} \sqrt{P_{I,n}}\, h_{I,n} x_{I,n}(t) + n_1(t) \tag{2}$$

where P_S is the transmit power at S, $P_{I,n}$ is the transmit power of nth-interference, $x_S(t)$ and $x_{I,n}(t)$ are the transmit signal from S and that from nth-interference, satisfying $\mathrm{E}[|x_S(t)|^2] = 1$ and $\mathrm{E}[|x_{I,n}(t)|^2] = 1$, $n_0(t)$ and $n_1(t)$ are the zero mean additive white Gaussian noises (AWGNs) at D and R, obeying $n_0(t) \sim \mathbb{N}_c(0,\sigma_0^2)$ and $n_1(t) \sim \mathbb{N}_c(0,\sigma_1^2)$, respectively, and E[.] denotes the expectation operator. In the second time slot, the received signal at R is amplified and forwarded to D through the fading channel h_2. Accordingly, the signal received at D can be written as

$$\begin{aligned} y_2(t) &= \sqrt{P_R}\, G_R h_2 y_1(t) + n_2(t) \\ &= \sqrt{P_S P_R}\, G_R h_1 h_2 x_S(t) + \sum_{n=1}^{N} \sqrt{P_{I,n} P_R}\, G_R h_{I,n} h_2 x_{I,n}(t) \\ &\quad + \sqrt{P_R}\, G_R h_2 n_1 + n_2(t) \end{aligned} \tag{3}$$

where P_R denotes the transmit power at R, $n_2(t)$ the AWGN at D with zero mean and variance σ_2^2. Furthermore, G_R is the amplify gain at R, given by [9]

$$G_R = \sqrt{\frac{1}{P_S |h_1|^2 + \sum_{n=1}^{N} P_{I,n} |h_{I,n}|^2 + \sigma_1^2}}\ . \tag{4}$$

Following the application of maximal ratio combining (MRC) at D, it is not difficult to find that the output SINR can be expressed as

$$\begin{aligned} \gamma &= \frac{P_S |h_0|^2}{\sigma_0^2} + \frac{P_S P_R G_R^2 |h_1|^2 |h_2|^2}{\sum_{n=1}^{N} P_{I,n} P_R G_R^2 |h_{I,n}|^2 |h_2|^2 + P_R G_R^2 |h_2|^2 \sigma_1^2 + \sigma_2^2} \\ &= \frac{P_S |h_0|^2}{\sigma_0^2} + \frac{\frac{P_S |h_1|^2}{\sigma_1^2}\frac{P_R |h_2|^2}{\sigma_2^2}}{\frac{P_S |h_1|^2}{\sigma_1^2} + \left(\frac{P_R |h_2|^2}{\sigma_2^2} + 1\right)\left(\sum_{n=1}^{N} \frac{P_{I,n} |h_{I,n}|^2}{\sigma_1^2} + 1\right)} \\ &= \gamma_0 + \frac{\gamma_1 \gamma_2}{\gamma_1 + (\gamma_2 + 1)(\gamma_3 + 1)} \\ &\cong \gamma_0 + \gamma_R \end{aligned} \tag{5}$$

where $\gamma_0 = |h_0|^2 \bar{\gamma}_0$, $\gamma_1 = |h_1|^2 \bar{\gamma}_1$, $\gamma_2 = |h_2|^2 \bar{\gamma}_2$ and $\gamma_3 = \sum_{n=1}^{N} |h_{I,n}|^2 \bar{\gamma}_{3,n}$ with $\bar{\gamma}_0 = P_S / \sigma_0^2$, $\bar{\gamma}_1 = P_S / \sigma_1^2$, $\bar{\gamma}_2 = P_R / \sigma_2^2$ and $\bar{\gamma}_{3,n} = P_{I,n} / \sigma_1^2$ being the average signal-to-noise ratios (SNRs). In (5), γ_0 and γ_R represent the output signal-to-noise ratio (SNR) of direct S-D link and the output SINR of the S-R-D relay link at D, respectively. Note that a similar expression has also been derived in [6, equation 5] for the case of ignoring direct link between S and D and noise at the relay, indicating that our work extends the study of previous works to a more general case.

3. Moment Generating Function

MGF is an important approach to evaluate the performance of wireless systems, which is defined as $M_r(s) = \mathrm{E}\left[e^{-s\gamma}\right]$. Clearly, the MGF of γ in (5) can be obtained as

$$\begin{aligned} M_r(s) &= \mathrm{E}\left[e^{-s(\gamma_0 + \gamma_R)}\right] = \mathrm{E}\left[e^{-s\gamma_0}\right]\mathrm{E}\left[e^{-s\gamma_R}\right] \\ &= M_{r_0}(s) M_{r_R}(s) \end{aligned} \tag{6}$$

where $M_{r_0}(s)$ and $M_{r_R}(s)$ are the MGF of γ_0 and γ_R, respectively. Since the *S-D* link undergoes Rayleigh fading, $|h_0|^2$ has a Chi-square distribution with 2 degrees of freedom and variance 1/2, the probability density function (PDF) is given by

$$f_{\gamma_0}(x) = \frac{1}{\bar{\gamma}_0} \exp\left(-\frac{x}{\bar{\gamma}_0}\right). \tag{7}$$

Thus, the MGF of γ_0 can be obtained as [4]

$$M_{\gamma_0}(s) = \mathrm{E}\left[e^{-s\gamma_0}\right] = \int_0^{\infty} e^{-sx} f_{\gamma_0}(x)\, dx = \frac{1}{1 + s\bar{\gamma}_0}\ . \tag{8}$$

To obtain the analytical MGF $M_{r_R}(s)$ of γ_R, we first express the upper bound of γ_R as

$$\tilde{\gamma}_R = \min\left(\frac{\gamma_1}{\gamma_3 + 1}, \gamma_2\right) \cong \min(\gamma_{13}, \gamma_2) \tag{9}$$

for mathematical tractability. Thus, the MGF of γ_R can be approximated as

$$M_{\gamma_R}(s) \approx M_{\tilde{\gamma}_R}(s) = \mathrm{E}\left[e^{-s\tilde{\gamma}_R}\right] = \int_0^\infty e^{-sx} f_{\tilde{\gamma}_R}(x)\,dx = \int_0^\infty s e^{-sx} F_{\tilde{\gamma}_R}(x)\,dx \quad (10)$$

where $F_{\tilde{\gamma}_R}(x)$ is the cumulative distribution function (CDF) of $\tilde{\gamma}_R$, given by

$$F_{\tilde{\gamma}_R}(x) = 1 - \left[1 - F_{\gamma_{13}}(x)\right]\left[1 - F_{\gamma_2}(x)\right] \quad (11)$$

with $F_{\gamma_{13}}(x)$ and $F_{\gamma_2}(x)$ being the CDF of γ_{13} and γ_2. Next, according to [12], the PDF of $\gamma_3 = \sum_{n=1}^{N} |h_{I,n}|^2 \bar{\gamma}_{3,n}$ can be written as

$$f_{\gamma_3}(x) = \sum_{i=1}^{t}\sum_{j=1}^{v_i} \frac{C_{i,j}}{(j-1)!\left(\bar{\gamma}_{3,i}\right)^j} x^{j-1} \exp\left(-\frac{x}{\bar{\gamma}_{3,i}}\right) \quad (12)$$

with the coefficient $C_{i,j}$ given by

$$C_{i,j} = \frac{1}{(v_i - j)!\,\bar{\gamma}_{3,i}^{v_i - j}} \frac{\partial^{v_i - j}}{\partial s^{v_i - j}} \left[\prod_{k=1,k\neq i}^{t_n} \left(\frac{1}{1 + s\bar{\gamma}_{3,k}}\right)^{v_k}\right]\Bigg|_{s=-\bar{\gamma}_{3,i}^{-1}} \quad (13)$$

where v_i denotes the repeated times of $\bar{\gamma}_{3,i}$ satisfying $\sum_{i=1}^{t} v_i = N$. Consider the *S-R* link experiences Rayleigh fading, the CDF of γ_1 can be written as

$$F_{\gamma_1}(x) = 1 - \exp\left(-\frac{x}{\bar{\gamma}_1}\right). \quad (14)$$

Therefore, $F_{\gamma_{13}}(x)$ can be calculated as

$$\begin{aligned} F_{\gamma_{13}}(x) &= \Pr\left(\frac{\gamma_1}{\gamma_3 + 1} \le x\right) = \int_0^\infty \int_0^{x(1+z)} f_{\gamma_1}(y) f_{\gamma_3}(z)\,dy\,dz \\ &= \int_0^\infty F_{\gamma_1}\left(x(1+z)\right) f_{\gamma_3}(z)\,dz \\ &= 1 - \exp\left(-\frac{x}{\bar{\gamma}_1}\right) \sum_{i=1}^{t}\sum_{j=1}^{v_i} \frac{C_{i,j}}{(j-1)!}\left(\frac{1}{\bar{\gamma}_{3,i}^j}\right)^j \\ &\times \int_0^\infty z^{j-1} \exp\left(-\left(\frac{z}{\bar{\gamma}_1} + \frac{1}{\bar{\gamma}_{3,i}}\right)\right) dz . \end{aligned} \quad (15)$$

With the help of identity [13, equation 3.351.3], we have

$$F_{\gamma_{13}}(x) = 1 - \exp\left(-\frac{x}{\bar{\gamma}_1}\right) \sum_{i=1}^{t}\sum_{j=1}^{v_i} C_{i,j} \left(1 + \frac{\bar{\gamma}_{3,i}}{\bar{\gamma}_1}\right)^{-j} . \quad (16)$$

Meanwhile, the *R-D* link is subject to Rayleigh fading, one can obtain

$$F_{\gamma_2}(x) = 1 - \exp\left(-\frac{x}{\bar{\gamma}_2}\right). \quad (17)$$

Substituting (16) and (17) into (11), one can obtain the closed-form expression of $F_{\tilde{\gamma}_R}(x)$ as

$$F_{\tilde{\gamma}_R}(x) = 1 - \exp\left(-\left(\frac{1}{\bar{\gamma}_1} + \frac{1}{\bar{\gamma}_2}\right)x\right) \sum_{i=1}^{t}\sum_{j=1}^{v} C_{i,j} \left(1 + \frac{\bar{\gamma}_{3,i}}{\bar{\gamma}_1}\right)^{-j} \quad (18)$$

Finally, by plugging (18) into (10), after some algebraic manipulations, the MGF of γ_R can be approximately calculated as

$$\begin{aligned} M_{\gamma_R}(s) &\approx 1 - s\sum_{i=1}^{t}\sum_{j=1}^{v} C_{i,j} \int_0^\infty \left(1 + \frac{\bar{\gamma}_{3,i}}{\bar{\gamma}_1}\right)^{-j} \exp\left(-\left(s + \frac{1}{\bar{\gamma}_1} + \frac{1}{\bar{\gamma}_2}\right)x\right) dx \\ &= 1 - s\sum_{i=1}^{t}\sum_{j=1}^{v_i} C_{i,j} \frac{\bar{\gamma}_1}{\bar{\gamma}_{3,i}} U\left(1, 2-j; \left(s + \frac{1}{\bar{\gamma}_1} + \frac{1}{\bar{\gamma}_2}\right)\frac{\bar{\gamma}_1}{\bar{\gamma}_{3,i}}\right) \end{aligned} \quad (19)$$

where $U(a, b; x)$ denotes the confluent hypergeometric function [14]. In deriving (19), we have applied [14, equation 2.3.6.9]. Furthermore, using (8) and (19) into (6), one can readily obtain the approximate expression of $M_r(s)$, giving

$$\begin{aligned} M_r(s) &\approx \frac{1}{1 + s\bar{\gamma}_0} - \frac{s}{1 + s\bar{\gamma}_0} \sum_{i=1}^{t}\sum_{j=1}^{v_i} C_{i,j} \frac{\bar{\gamma}_1}{\bar{\gamma}_{3,i}} \\ &\times U\left(1, 2-j; \left(s + \frac{1}{\bar{\gamma}_1} + \frac{1}{\bar{\gamma}_2}\right)\frac{\bar{\gamma}_1}{\bar{\gamma}_{3,i}}\right). \end{aligned} \quad (20)$$

4. Average Symbol Error Rate

The ASER of wireless communication systems with various modulation formants over fading channels can be expressed as [15]

$$P_s = \int_0^\theta a M_\gamma\left(\frac{b}{\sin^2\phi}\right) d\phi \quad (21)$$

where a, b and θ are the modulation specific constants. However, to the best of our knowledge, due to the integral, the closed-form solution is mathematically unavailable by directly employing the above formula. To overcome this, we provide an approximate yet accurate alternative method to evaluate the ASER.

For the case of M-ary phase-shift keying (M-PSK) modulation signal, we have

$$\begin{aligned} P_{M-\mathrm{PSK}} &= \frac{1}{\pi}\int_0^{(M-1)\pi/M} M_\gamma\left(\frac{\sin^2(\pi/M)}{\sin^2\phi}\right) d\phi \\ &= \underbrace{\frac{1}{\pi}\int_0^{\pi/2} M_\gamma\left(\frac{\sin^2(\pi/M)}{\sin^2\phi}\right) d\phi}_{I_1} + \underbrace{\frac{1}{\pi}\int_{\pi/2}^{(M-1)\pi/M} M_\gamma\left(\frac{\sin^2(\pi/M)}{\sin^2\phi}\right) d\phi}_{I_2} \end{aligned} \quad (22)$$

By using the definition of MGF, L_1 can be calculated as

$$L_1 = \frac{1}{\pi}\int_0^{\pi/2}\int_0^{\infty}\exp\left(-\frac{x\sin^2(\pi/M)}{\sin^2\phi}\right)f_\gamma(x)\,dx\,d\phi$$
$$\approx \int_0^{\infty}\left(\frac{1}{12}\exp\left(-x\sin^2\left(\frac{\pi}{M}\right)\right)\right.$$
$$\left.+\frac{1}{4}\exp\left(-\frac{4x\sin^2(\pi/M)}{3}\right)\right)f_\gamma(x)\,dx$$
$$=\frac{1}{12}M_\gamma\left(\sin^2\left(\frac{\pi}{M}\right)\right)+\frac{1}{4}M_\gamma\left(\frac{4}{3}\sin^2\left(\frac{\pi}{M}\right)\right). \quad (23)$$

In deriving (23), we have applied the following approximation [16]

$$\frac{2}{\pi}\int_0^{\frac{\pi}{2}}\exp\left(-\frac{x}{\sin^2\theta}\right)d\theta \approx \frac{1}{6}\exp(-x)+\frac{1}{2}\exp\left(-\frac{4}{3}x\right). \quad (24)$$

Similarly, with the help of identity [16]

$$\frac{1}{\pi}\int_{\frac{\pi}{2}}^{\frac{(M-1)}{M}\pi}\exp\left(-\frac{x}{\sin^2\theta}\right)d\theta \approx \left(\frac{(M-1)}{2M}-\frac{1}{4}\right)$$
$$\times\left(\exp(-x)+\exp\left(-\frac{x}{\sin^2((M-1)\pi/M)}\right)\right) \quad (25)$$

L_2 can be approximately obtained as

$$L_2 = \frac{1}{\pi}\int_{\pi/2}^{(M-1)\pi/M}\int_0^{\infty}\exp\left(-\frac{x\sin^2(\pi/M)}{\sin^2\phi}\right)f_\gamma(x)\,dx\,d\phi$$
$$\approx\int_0^{\infty}\left(\left(\frac{M-1}{2M}-\frac{1}{4}\right)\left(\exp\left(-x\sin^2\left(\frac{\pi}{M}\right)\right)\right.\right.$$
$$\left.\left.+\exp\left(-\frac{x\sin^2(\pi/M)}{\sin^2((M-1)\pi/M)}\right)\right)\right)f_\gamma(x)\,dx$$
$$=\left(\frac{M-1}{2M}-\frac{1}{4}\right)M_\gamma\left(\sin^2\left(\frac{\pi}{M}\right)\right)$$
$$+\left(\frac{M-1}{2M}-\frac{1}{4}\right)M_\gamma\left(\frac{\sin^2(\pi/M)}{\sin^2((M-1)\pi/M)}\right). \quad (26)$$

Furthermore, the approximate ASER expression for M-PSK modulation signal can be written as

$$P_{M-\mathrm{PSK}} \approx \left(\frac{M-1}{2M}-\frac{1}{4}\right)M_\gamma\left(\frac{\sin^2(\pi/M)}{\sin^2((M-1)\pi/M)}\right)$$
$$+\left(\frac{M-1}{2M}-\frac{1}{6}\right)M_\gamma\left(\sin^2\left(\frac{\pi}{M}\right)\right)+\frac{1}{4}M_\gamma\left(\frac{4}{3}\sin^2\left(\frac{\pi}{M}\right)\right). \quad (27)$$

In a similar manner, the ASERs of M-ary Pulse Amplitude Modulation (M-PAM) signals can be calculated as

$$P_{M-\mathrm{PAM}} = 2\left(\frac{M-1}{\pi M}\right)\int_0^{\frac{\pi}{2}}M_\gamma\left(\frac{3}{(M^2-1)\sin^2\phi}\right)d\phi$$
$$=2\left(\frac{M-1}{\pi M}\right)\int_0^{\frac{\pi}{2}}\int_0^{\infty}\exp\left(-\frac{3x}{(M^2-1)\sin^2\phi}\right)f_\gamma(x)\,dx\,d\phi$$
$$\approx\left(\frac{M-1}{\pi M}\right)\int_0^{\infty}\left(\frac{1}{3}\exp\left(-\frac{3x}{M^2-1}\right)+\exp\left(-\frac{4x}{3(M^2-1)}\right)\right)f_\gamma(x)\,dx$$
$$=\frac{M-1}{M}\left(\frac{1}{6}M_\gamma\left(\frac{3}{M^2-1}\right)+\frac{1}{2}M_\gamma\left(\frac{4}{M^2-1}\right)\right). \quad (28)$$

As for M-ary Quadrature Amplitude Modulation (M-QAM) signals, it can be considered as two independent $\sqrt{M}$ -PAM signals with its ASER being expressed as

$$P_{M-\mathrm{QAM}} = \underbrace{\frac{4}{\pi}\left(1-\frac{1}{\sqrt{M}}\right)\int_0^{\pi/2}M_\gamma\left(\frac{3}{(M-1)\sin^2\phi}\right)d\phi}_{L_3}$$
$$\underbrace{-\frac{4}{\pi}\left(1-\frac{1}{\sqrt{M}}\right)^2\int_0^{\pi/4}M_\gamma\left(\frac{3}{(M-1)\sin^2\phi}\right)d\phi}_{L_4}. \quad (29)$$

Following a similar manner to derive (23), L_3 can be obtained as

$$L_3 \approx \left(1-\frac{1}{\sqrt{M}}\right)\left(\frac{1}{3}M_\gamma\left(\frac{3}{M-1}\right)+M_\gamma\left(\frac{4}{M-1}\right)\right). \quad (30)$$

In addition, by applying an approximate expression [3], we have

$$L_4 \approx \frac{1}{6}\left(1-\frac{1}{\sqrt{M}}\right)^2\left(3M_\gamma\left(\frac{6}{M-1}\right)+M_\gamma\left(\frac{3}{M-1}\right)\right). \quad (31)$$

Therefore, an approximate expression of $P_{M-\mathrm{QAM}}(\gamma)$ can be calculated as

$$P_{M-\mathrm{QAM}}(\gamma) \approx \left(1-\frac{1}{\sqrt{M}}\right)\left(\frac{1}{3}M_\gamma\left(\frac{3}{M-1}\right)+M_\gamma\left(\frac{4}{M-1}\right)\right)$$
$$+\frac{1}{6}\left(1-\frac{1}{\sqrt{M}}\right)^2\left(3M_\gamma\left(\frac{6}{M-1}\right)+M_\gamma\left(\frac{3}{M-1}\right)\right). \quad (32)$$

5. Numerical Results

In this section, we confirm the validity of the derived analytical results through comparison with Monte Carlo simulations, and investigate the impacts of the direct link and CCIs on the performance of the considered cooperative relay network. Here we assume that R is corrupted by 3 CCIs with the SNRs denoted as $\bar{\gamma}_{I,1}$, $\bar{\gamma}_{I,2} = 0.6\bar{\gamma}_{I,1}$ and $\bar{\gamma}_{I,3} = 0.8\bar{\gamma}_{I,1}$. In all the plots, we define $\bar{\gamma}_1 = \bar{\gamma}_2 = \bar{\gamma}$,

$\eta = \overline{\gamma}/\overline{\gamma}_0$ and $\rho = \overline{\gamma}/\overline{\gamma}_{I,1}$, where the cases of no direct link and no interferences are denoted as $\eta = \infty$ dB and $\rho = \infty$ dB.

Fig. 2, Fig. 3 and Fig. 4 show, respectively, the ASER versus $\overline{\gamma}$ in terms of QPSK, 8PAM and 16QAM modulation for different η. As we see, since the approximate method is used to calculate the MGF of the output SINR in (10), there exist a little deviation at low SNR, the good match between the analytical results and the Monte Carlo experiments in general can still be obtained. The scenarios confirm the validity of the derived analytical expressions. . In addition, it can also be found that the ASER performance is improved with the increase of direct *S-D* link strength. This is because the direct link can enhance the output SINR.

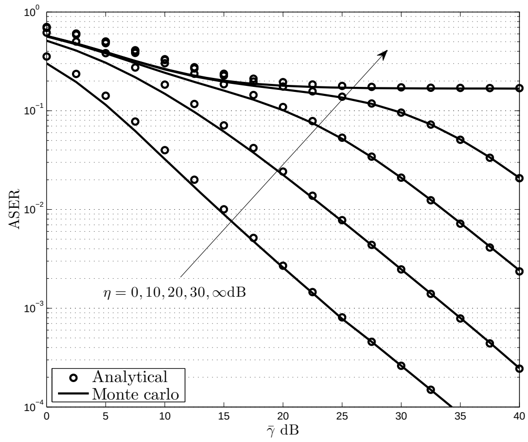

Fig. 2. ASER versus average SNR for QPSK with $N = 3$ and $\rho = 10$ dB.

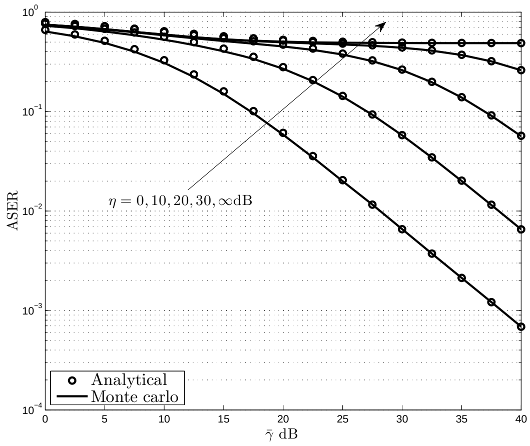

Fig. 3. ASER versus average SNR for 8PAM with $N = 3$ and $\rho = 10$ dB.

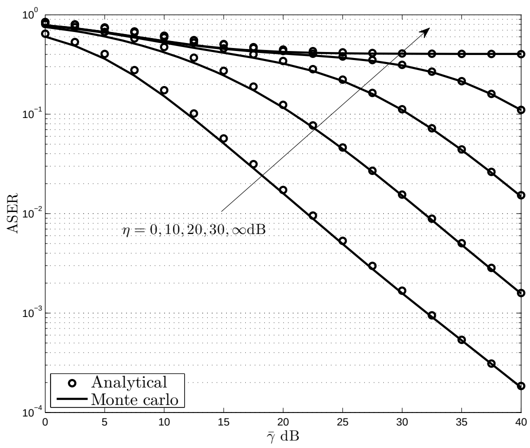

Fig. 4. ASER versus average SNR for 16QAM with $N = 3$ and $\rho = 10$ dB.

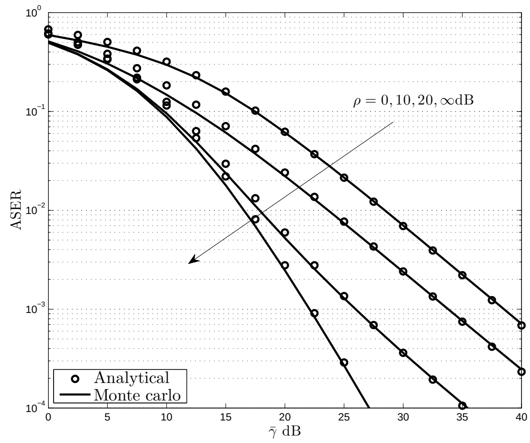

Fig. 5. ASER versus average SNR for QPSK with $N = 3$ and $\eta = 10$ dB.

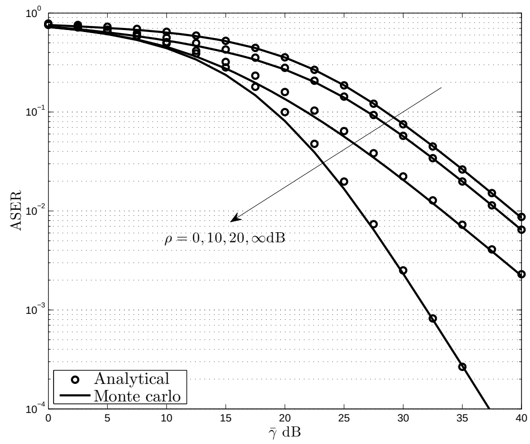

Fig. 6. ASER versus average SNR for 8PAM with $N = 3$ and $\eta = 10$ dB.

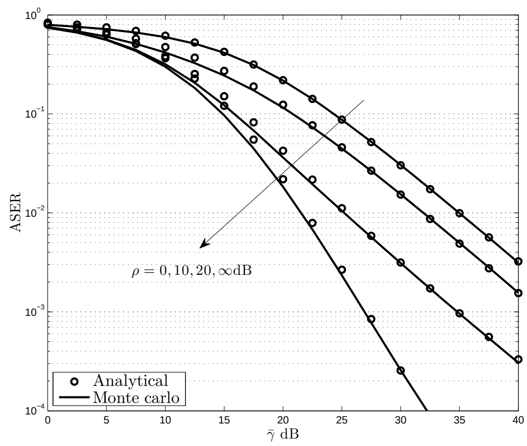

Fig. 7. ASER versus average SNR for 16QAM with $N = 3$ and $\eta = 10$ dB.

Fig. 5, Fig. 6 and Fig. 7 depict the ASERs of QPSK, 8PAM and 16QAM modulations against SNR for various ρ. As we expect, at middle and high SNR regimes, a well agreement between the analytical results and Monte Carlo experiments is observed. Furthermore, it can also be seen that the ASERs of the network decreases with the increase of ρ. This is because when ρ is increased, the CCIs are degraded, resulting in the increase of output instantaneous SNR at the destination.

6. Conclusions

In this paper, a dual-hop AF cooperative relay network has been investigated in the presence of *S-D* direct link and multiple CCIs at the relay. In particular, we have derived the new analytical expressions for the MGF of the output SINR as well as the ASER of the system. Furthermore, we have also provided numerical results to demonstrate the validity of the theoretical analysis, and examined the effects of direct link and interference on the considered AF relay network. It should be pointed out the extension of the presented method to the more complex networks, such as more than one nodes having multi-antenna and/or multiple relays are exploited to amplify the signals, is our further work.

Acknowledgements

This work is supported by the National Natural Science Foundation of China (61271255), the Natural Science Foundation of Jiangsu Province (BK20131068), and the Open Research Fund of National Mobile Communications Research Laboratory in Southeast University (2012D15).

References

[1] KAKITANI, M., BRANTE, G., SOUZA, R. Energy efficiency analysis of a two-dimensional cooperative wireless sensor network with relay selection. *Radioengineering,* 2013, vol. 22, no. 2, p. 549 to 557.

[2] FAN, Z. J., GUO, D. X., ZHANG, B. N. Outage probability and power allocation for two-way DF relay networks with relay selection. *Radioengineering,* 2012, vol. 21, no. 3, p. 795–801.

[3] LI, M., LIN, M., YU, Q., ZHU, W.-P., DONG, L. Optimal beamformer design for two hop MIMO AF relay networks over Rayleigh fading channels. *IEEE Journal on Selected Areas in Communications,* 2012, vol. 30, no. 8, p. 1402–1414.

[4] OUYANG, J., LIN, M., ZHUANG, Y. Performance analysis of cooperative relay networks over asymmetric fading channels. *Electronics Letters*, 2012, vol. 48, no. 21, p.1370–1371.

[5] LI, M., AN, K., OUYANG, J., WEI, H., et al. Effect of feedback delay on dual-hop fixed gain relay networks over mixed fading channels. *Transactions on Emerging Telecommunications Technologies*, 2014, vol. 25, no. 10, p. 1045–1055.

[6] SURAWEERA, H. A., HARI, K. G., NALLANATHAN, A. Performance analysis of two hop amplify-and-forward systems with interference at the relay. *IEEE Communications Letters*, 2010, vol. 14, no. 8, p. 692–694.

[7] TRIGUI, I., AFFES, S., STEPHENNE, A. On the ergodic capacity of amplify-and-forward relay channels with interference in Nakagami-m fading. *IEEE Transactions on Communications*, 2013, vol. 61, no. 8, p. 3136–3145.

[8] AN, K., LIN, M., OUYANG, J., WEI, H. Beamforming in dual-hop AF relaying with imperfect CSI and co-channel interference. *Wireless Personal Communication*, 2014, vol. 78, no. 2, p. 1187 to 1197.

[9] PHAN, H., DUONG, T. Q., ELKASHLAN, M., ZEPERNICK, H.-J. Beamforming amplify-and-forward relay networks with feedback delay and interference. *IEEE Signal Processing Letters*, 2012, vol. 19, no. 1, p. 16–19.

[10] AN, K., LIN, M., OUYANG, J., WEI, H. Performance analysis of beamforming in two-hop AF relay networks with antenna correlation and interference. *AEU-International Journal of Electronics and Communications*, 2014, vol. 68, no. 7, p. 587–594.

[11] DING, H., HE, C., JIANG, L.-G. Performance analysis of fixed gain MIMO relay systems in presence of co-channel interference. *IEEE Communications Letters*, 2012, vol. 16, no. 7, p. 1133–1136.

[12] CUI, X. W., FENG, Z. M. Lower capacity bound for MIMO correlated fading channels with keyhole. *IEEE Communications Letters*, 2004, vol. 8, p. 500–502.

[13] GRADSHTEYN, I. S., RYZHIK, I. M. *Table of Integrals, Series, and Products*. 7th ed. Academic Press, 2007.

[14] PRUDNIKOV, A. P., BRYCHKOV, Y. A., MARICHEV, O. I. *Integrals and Series*. Vol. 1, 1st ed. New York (USA): Gordon and Breach Science Publishers, 1986.

[15] SIMON, M. K., ALOUINI, M. S. *Digital Communication over Fading Channels*. 2nd ed. Wiley, 2005.

[16] MCKAY, M. R., ZANELLA, A., COLLINGS, I. B., et al. Error probability and SINR analysis of optimum combining in Rician fading. *IEEE Transactions on Communications*, 2009, vol. 57, no. 3, p. 676–687.

About Authors ...

Min LIN was born in Zhejiang Province, China, in 1972. He received the B.S. degree from the National University of Defense Technology, the M.S. degree from Nanjing Institute of Communication Engineering, and Ph.D. degree from the Southeast University in 1993, 2000, and 2008, respectively, all in Electrical Engineering. His current research interests include wireless communications, array signal processing and antenna technology. He is the author or coauthor of more than 70 papers, and holds 4 patents. Dr. Lin is a Member of IEEE and Senior Member of Chinese Institute of Electronics (CIE).

Heng WEI (corresponding author) was born in Jiangsu Province, China, in 1989. He received the B.S. degree from the College of Electronic and Information Engineering, Nanjing University of Aeronautics and Astronautics in 2012. He is currently working towards the M.S. degree in Electronic Engineering, from the PLA University of Science and Technology. His research interests are MIMO technique and wireless relay communications.

Jian OUYANG was born in Jiangsu Province, China, in 1983. He received the B.S., M.S. and Ph.D. degrees from the College of Computer Science and Technology, Nanjing University of Aeronautics and Astronautics in 2007, 2010, and 2014 respectively. Since July 2014, he has been with the Institute of Signal Processing and Transmission, Nanjing University of Posts and Telecommunications (NUPT). His current research interests include wireless cooperative networks.

Kang AN was born in Xinjiang Uygur autonomous region, China, in 1989. He received the B.S. degree from the College of Electronic and Information Engineering, Nanjing University of Aeronautics and Astronautics in 2011, and M.S. degree in Electronic Engineering, from the College of Communications Engineering, PLA University of Science and Technology in 2014. He is currently working towards the Ph.D. degree at the College of Communications Engineering, PLA University of Science and Technology. His research interests include hybrid satellite-terrestrial network, cooperative communications, cognitive radio and array signal processing.

Can YUAN was born in Anhui Province, China, in 1991. He received the B.S. degree from the College of Science, Nanjing University of Aeronautics and Astronautics in 2013. He is currently working towards the M.S. degree in Communication Engineering at the College of Communications Engineering, PLA University of Science and Technology. His research interests are array signal processing and wireless cooperative communications.