UAV Relay Assisted Cooperative Jamming for Covert Communications over Rician Fading

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***Abstract*—Covert communication can hide the legitimate transmission from unauthorized eavesdropping. Benefiting from the deployment flexibility, unmanned aerial vehicles (UAVs) can be utilized to enhance communication confidentiality. In this correspondence, we consider a covert communication network with the aid of a full-duplex UAV relay, which is employed to help the transmission and confuse the warden. The warden adopts a radiometer to detect the covert transmission. We first find the optimal detection threshold and calculate the minimum detection error probability. Furthermore, a closed-form expression of outage probability via UAV relaying is derived over Rician fading. Then, a power optimization problem is formulated to maximize the effective convert throughput with covertness constraint. Numerical results illustrate that the cooperative jamming can disrupt the warden, and the optimal power tradeoff can guarantee the covert transmission effectively.**

***Index Terms*—Cooperative jamming, covert communication, power allocation, rician fading, UAV relaying.**

I. Introduction

With the explosive growth of wireless services, communication security has become a crucial requirement for wireless networks. Especially concealing the transmission of secret information is essential for achieving communication confidentiality. Covert communication, as an emerging security technique, can prevent the transmission from being intercepted [1]. The pioneering work by Bash *et al.* in [2] has proved the square root law that at most $\mathcal{O}(\sqrt{n})$ bits can be transmitted over n channel uses covertly. The work by Sobers *et al.* in [3] exploited the uncertainties generated by an uninformed jammer to achieve covertness.

Recently, plenty of works have focused on the covert design for different scenarios, such as single-hop [4], [5] and two-hop networks [6], [7]. In [4], a full-duplex receiver was utilized by Shahzad *et al.* to generate artificial noise (AN) to confuse the warden while receiving signal from the transmitter. Shu *et al.* in [5] proved that transmitting AN with fixed power can improve the performance of covert communication under finite blocklength. However, single-hop networks usually require higher transmit power to mitigate the effects of channel fading. Hu *et al.* proposed two schemes in [6] that adopt rate control or power control to help a greedy relay transmit the covert information. In [7], Sun *et al.* explored the optimal power control of relay under full-duplex and half-duplex modes to maximize the covert rate.

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It is worth noting that unmanned aerial vehicle (UAV) is very suitable for covert communications due to its mobility and deployment flexibility [8]. However, the line-of-sight (LoS) channels in UAV communication are also vulnerable to malicious warden [9]. To solve this problem, there are some studies investigating the application of UAV in covert communications [10]–[12]. In [10], a joint UAV trajectory and transmit power optimization was proposed by Zhou *et al.* to maximize the average covert transmission rate. Yan *et al.* in [11] further explored the optimal location of UAV to maximize the covert communication performance subject to a delay constraint. Chen *et al.* proposed a UAV-relayed covert communication scheme against a flying warden in [12].

However, to our best knowledge, only a few works have investigated the UAV as a relay for covert communications [12]. In addition, most of the existing works on UAV covert communications have adopted a simplified LoS channel. This motivates us to study the performance of UAV relaying in covert communication by employing more realistic and accurate Rician fading. As a consequence, this correspondence considers a UAV relay assisted wireless network, where the full-duplex relay not only receives the signal, but also transmits AN to confuse the warden to achieve covert communication over Rician fading. The main contributions are summarized as follows. First, the optimal threshold of the warden and the conditions under which the warden makes detection errors are derived. Then, we present the closed-form expression of outage probability over Rician fading, and define the effective covert rate as the objective function. Finally, we present the power optimization for source and relay to maximize the covert rate subject to the covertness constraint.

II. System Model

Consider a one-way relaying system in Fig. 1, where the source ($\mathcal{S}$) attempts to transmit to the destination ($\mathcal{D}$) with the help of a UAV relay ($\mathcal{R}$). The direct link between $\mathcal{S}$ and $\mathcal{D}$ is unavailable due to the blocking. The UAV is hovering with a quasi-stationary state during the whole relaying. At the same time, a warden ($\mathcal{W}$) near $\mathcal{S}$ tries to make a critical decision about whether the transmission is on or off. Assume that $\mathcal{W}$ knows the potential time when $\mathcal{S}$ may transmit, and it has complete knowledge about the network structure.

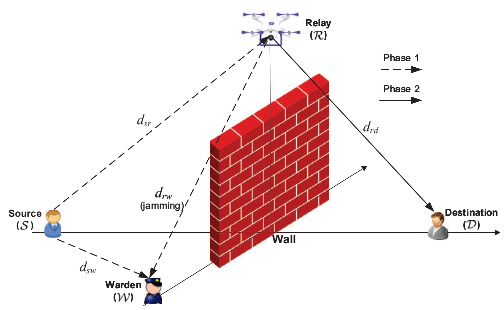

Fig. 1. UAV relay system for covert communication.

All the wireless channels are assumed to undergo both large-scale path-loss and small-scale fading, and the channel coefficients between $\mathcal{S}$, $\mathcal{W}$, $\mathcal{D}$ and $\mathcal{R}$ can be modeled as

$$h_{ij} = \sqrt{\beta_{ij}} g_{ij}, \;\; ij \in \{sr, rd, rw, sw\}, \tag{1}$$

where $\beta_{ij} = \beta_0 d_{ij}^{-\alpha_{ij}}$ denotes the large-scale average channel power, β_0 is the channel power gain at unit distance, and d_{ij} represents the distance between any two given nodes. Besides, the path-loss exponent α_{ij} can be calculated by using the angle-dependent probability model as

$$\alpha_{ij} = \xi_1(1 - P_{\mathrm{LoS}}(\theta_{ij})) + \xi_0 = \xi_1\left(1 - \frac{1}{1 + a\mathrm{e}^{-b(\theta_{ij}-a)}}\right) + \xi_0, \tag{2}$$

where a and b are determined by the environmental characteristics, ξ_1 and ξ_0 denote the basic path-loss coefficients, and the elevation angle $\theta_{ij} = \arcsin(\frac{H}{d_{ij}})$. H denotes the UAV height. The small-scale Rician fading coefficient g_{ij} with $\mathbb{E}[|g_{ij}|^2] = 1$ can be modeled as [13]

$$g_{ij} = \sqrt{\frac{K_{ij}}{K_{ij}+1}} g + \sqrt{\frac{1}{K_{ij}+1}} \tilde{g}, \;\; ij \in \{sr, rd, rw\}, \tag{3}$$

where g is the deterministic line-of-sight (LoS) channel component with $|g| = 1$, and $\tilde{g}$ is a zero-mean unit-variance circularly symmetric complex Gaussian (CSCG) random variable of the scattered component. The Rician factor K_{ij} is a function of the elevation angle θ_{ij}, which can be given by

$$K_{ij} = \kappa_0 + (\kappa_{\frac{\pi}{2}} - \kappa_0) 2\theta_{ij}/\pi, \tag{4}$$

where κ_0 and $\kappa_{\frac{\pi}{2}}$ are the environment depended parameters. It can be observed that Rician fading can be reduced to Rayleigh fading when $K = 0$, which is used to describe the small-scale fading of terrestrial channel as $g_{sw} \sim \mathcal{CN}(0, 1)$. Similarly, the path-loss exponent of terrestrial link can be calculated by substituting $\theta_{sw} = 0$ into (2).

III. Cooperative Relaying Scheme

Assume that $\mathcal{R}$ is equipped with both a receive antenna and a transmit antenna, and employs the amplify-and-forward (AF) model. We propose a cooperative UAV relaying scheme where $\mathcal{R}$ adopts the full-duplex mode for receiving and jamming in the first phase. The jamming signal can degrade the detection ability of $\mathcal{W}$. Then, the retransmission in the second phase enables the message of $\mathcal{S}$ to be received at $\mathcal{D}$ covertly.

In the first phase, $\mathcal{S}$ chooses to transmit with a fixed power P_s or nothing, and these two events are denoted by H_1 and H_0, respectively. When H_1 holds, $\mathcal{S}$ maps its messages to a sequence of n complex symbols $\mathbf{x}_s = [x_s(1), x_s(2), \ldots x_s(n)]$. Regardless of whether $\mathcal{S}$ sends the messages, $\mathcal{R}$ always transmits the jamming signal $\mathbf{x}_j = [x_j(1), x_j(2), \ldots x_j(n)]$ to confuse $\mathcal{W}$, with the transmit power P_r. The symbols are assumed to be independent identically distribution (i.i.d), satisfying $\mathbb{E}[x_s(i)x_s^\dagger(i)] = 1$ and $\mathbb{E}[x_j(i)x_j^\dagger(i)] = 1$, where $i = 1, \ldots, n$ represents the symbol index. It is worth noting that the self-interference generated by the full-duplex relay can be partially eliminated via the effective self-interference cancellation. The residual self-interference is denoted by $r(i)$, where $r(i) \sim \mathcal{CN}(0, \xi P_r \sigma^2)$ with the factor $0 < \xi < 1$. Thus, the received signal at $\mathcal{R}$ can be expressed as

$$y_r(i) = \sqrt{P_s \beta_{sr}} g_{sr} x_s(i) + r(i) + n_r(i), i = 1, \ldots, n, \tag{5}$$

where $n_r(i)$ is the noise at $\mathcal{R}$ with σ_r^2 as its variance, i.e., $n_r(i) \sim \mathcal{CN}(0, \sigma_r^2)$. For simplicity, we assume that $\sigma_k^2 = \sigma^2$ for $k \in \{r, w, d\}$.

In the second phase, $\mathcal{R}$ amplifies the received signal $y_r(i)$ with the coefficient G when H_1 holds. Automatic gain control is employed by $\mathcal{R}$ to ensure that the transmit power is P_r, where the coefficient can be expressed as $G = P_r/(P_s\beta_{sr}|g_{sr}|^2 + \xi P_r \sigma^2 + \sigma^2)$. We assume that even when $\mathcal{S}$ keeps silent, $\mathcal{R}$ also sends the jamming signal with P_r. The jamming codebook is known at $\mathcal{D}$, which can be successfully decoded and removed.

A. Requirement of Covert Communication

Since the UAV relay always transmits in Phase II, either the relayed signal or the jamming signal, we mainly focus on the testing of Phase I, and $\mathcal{W}$ has to make a decision whether $\mathcal{S}$ is transmitting in Phase I based on the received signal. There are two error probabilities for a detection device, i.e., miss detection $\mathbb{P}_{\mathrm{MD}}$ and false alarm $\mathbb{P}_{\mathrm{FA}}$. Accordingly, the covert communication can be achieved when the following constraint is satisfied. For any sufficiently small constant $\epsilon_c > 0$, the communication scheme should ensure the probability of detection error $\mathbb{P}_{\mathrm{E}} \triangleq \mathbb{P}_{\mathrm{FA}} + \mathbb{P}_{\mathrm{MD}} \geq 1 - \epsilon_c$ as n is assumed to be infinitely large, where ϵ_c indicates the covert communication requirement. From the perspective of $\mathcal{W}$, any kind of detection error made by $\mathcal{W}$ is intolerable.

B. Problem Formulation

In this work, we focus on maximizing the effective covert rate, which can be defined as

$$R = (1 - \mathbb{P}_{\mathrm{out}}) R_{sd}, \tag{6}$$

where R_{sd} represents the required transmission rate from $\mathcal{S}$ to $\mathcal{D}$, and $\mathbb{P}_{\mathrm{out}}$ denotes the outage probability of the UAV relaying, which will be detailed in Section V-A.

We propose a feasible and effective power allocation scheme for the relaying system to achieve the maximal covert throughput. Assume that the total transmit power $P_{\mathrm{sum}} = P_s + P_r$ is fixed, and set a factor ρ to define $P_s = \rho P_{\mathrm{sum}}$ and $P_r = (1-\rho)P_{\mathrm{sum}}, (0 < \rho < 1)$. Moreover, we should have a constraint to guarantee covertness. Thus, the optimization problem can be formulated as

$$(\mathrm{P1}): \max_{\rho} \;\; (1 - \mathbb{P}_{\mathrm{out}}) R_{sd} \tag{7a}$$

$$s.t. \;\; \mathbb{P}_{\mathrm{E}} \geq 1 - \epsilon_c, \tag{7b}$$

$$0 < \rho < 1. \tag{7c}$$

IV. Performance Analysis of Covertness

In this section, we focus on handling the covertness constraint. The expression of the detection error probability is first derived, and then we calculate the optimal detection threshold to minimize the average detection error probability under a worst-case scenario.

A. Detection Error Probability

The signal received at $\mathcal{W}$ for each symbol index under the hypothesis H_0 and H_1 can be given by

$$y_w(i)=\begin{cases}\Psi+n_w(i), & H_0\\ \sqrt{P_s\beta_{sw}}g_{sw}x_s(i)+\Psi+n_w(i), & H_1\end{cases} \quad (8)$$

where $\Psi=\sqrt{P_r\beta_{rw}}g_{rw}x_j(i)$ and $n_w(i)\sim\mathcal{CN}(0,\sigma^2)$. Note that the codebook for the signal x_s is usually unknown at $\mathcal{W}$.

In order to detect the transmission from $\mathcal{S}$, $\mathcal{W}$ uses the commonly-used likelihood ratio test and adopts a radiometer to conduct the binary detection [3]. Based on (8), $\mathcal{W}$ can take the average received power P_w as the test statistic, and the decision rule can be expressed as

$$P_w \triangleq \frac{1}{n}\sum_{i=1}^{n}|y_w(i)|^2 \underset{H_0}{\overset{H_1}{\gtrless}} \nu, \quad (9)$$

where ν denotes the detection threshold. Based on the strong law of large numbers, the probabilities of false alarm and miss detection at $\mathcal{W}$ are given in the following lemma.

Lemma 1: Let $\bar{\lambda}_1=P_r\beta_{rw}$ and $\bar{\lambda}_2=P_s\beta_{sw}$ denote the average received signal power of the $\mathcal{R}$-$\mathcal{W}$ and $\mathcal{S}$-$\mathcal{W}$ links, respectively. Let $c_1=\frac{K_{rw}+1}{\bar{\lambda}_1}-\frac{1}{\bar{\lambda}_2}>0$ and $c_2=\frac{K_{rw}(K_{rw}+1)}{c_1\bar{\lambda}_1}$ denote the constant coefficients. Then we have

$$\mathbb{P}_{\mathrm{FA}}=\begin{cases}1, & \nu<\sigma^2,\\ \Gamma, & \nu\geq\sigma^2,\end{cases} \quad (10)$$

and

$$\mathbb{P}_{\mathrm{MD}}=\begin{cases}0, & \nu<\sigma^2,\\ 1-\Gamma-\Theta, & \nu\geq\sigma^2,\end{cases} \quad (11)$$

where

$$\Gamma=Q_1\left(\sqrt{2K_{rw}},\sqrt{2(1+K_{rw})\frac{\nu-\sigma^2}{P_r\beta_{rw}}}\right). \quad (12)$$

$Q_1(\cdot)$ is the first order Marcum Q-function and

$$\Theta=\frac{(K_{rw}+1)e^{c_2}}{c_1\bar{\lambda}_1e^{K_{rw}}}e^{-\frac{\nu-\sigma^2}{\bar{\lambda}_2}}\Big(1-Q_1\Big(\sqrt{2c_2},\sqrt{2c_1(\nu-\sigma^2)}\Big)\Big). \quad (13)$$

Proof: According to the decision rule (9) and the strong law of large numbers, the probability of false alarm can be computed as

$$\begin{aligned}\mathbb{P}_{\mathrm{FA}}&=\Pr\left(P_w\geq\nu|H_0\right)=\Pr\left(P_r\beta_{rw}|g_{rw}|^2+\sigma^2\geq\nu|H_0\right),\\ &\overset{(a)}{=}\begin{cases}1, & \nu<\sigma^2,\\ Q_1\left(\sqrt{2K_{rw}},\sqrt{2(1+K_{rw})\frac{\nu-\sigma^2}{P_r\beta_{rw}}}\right), & \nu\geq\sigma^2,\end{cases}\end{aligned} \quad (14)$$

where (a) follows the cumulative distribution function (CDF) of $|g_{rw}|^2$. $Q_1(\cdot)$ is the first-order Marcum Q-function, and K_{rw} is the Rician K-factor between $\mathcal{R}$ and $\mathcal{W}$.

Similarly, the probability of miss detection can be given by as

$$\mathbb{P}_{\mathrm{MD}}=\Pr\left(P_r\beta_{rw}|g_{rw}|^2+P_s\beta_{sw}|g_{sw}|^2+\sigma^2\leq\nu\right). \quad (15)$$

Note that $\mathbb{P}_{\mathrm{MD}}$ involves the joint distribution of two independent random variables. Recalling the channel distribution in Section II, the random variable $\lambda_1=P_r\beta_{rw}|g_{rw}|^2$ follows a non-central chi-square distribution with the Rician factor K_{rw}. For convenience, we drop the subscript rw from K_{rw} to simplify the notation, and the probability density function (PDF) can be given by

$$f_{\lambda_1}(x)=\frac{K+1}{\bar{\lambda}_1e^K}e^{-\frac{K+1}{\bar{\lambda}_1}x}I_0\left(2\sqrt{\frac{K(K+1)}{\bar{\lambda}_1}x}\right), x\geq0, \quad (16)$$

where $\bar{\lambda}_1=P_r\beta_{rw}$ and $I_0(\cdot)$ is the first-kind zero-order modified Bessel function.

The second random variable $\lambda_2=P_s\beta_{sw}|g_{sw}|^2$ is exponentially distributed with the PDF given by

$$f_{\lambda_2}(x)=\frac{1}{\bar{\lambda}_2}e^{-x/\bar{\lambda}_2}, x\geq0, \quad (17)$$

where $\bar{\lambda}_2=P_s\beta_{sw}$. After some algebraic manipulations, the CDF of $\lambda_1+\lambda_2$ can be written as

$$\begin{aligned}F_\lambda(z)&=\Pr(\lambda_1+\lambda_2\leq z)=\int_0^z f_{\lambda_2}(\lambda_1+\lambda_2\leq z|\lambda_1)f_{\lambda_1}(\lambda_1)\mathrm{d}\lambda_1\\ &=\int_0^z(1-e^{\frac{\lambda_1-z}{\bar{\lambda}_2}})\frac{K+1}{\bar{\lambda}_1e^K}e^{-\frac{(K+1)\lambda_1}{\bar{\lambda}_1}}I_0\left(2\sqrt{\frac{K(K+1)\lambda_1}{\bar{\lambda}_1}}\right)\mathrm{d}\lambda_1\\ &=\Lambda(z)-\Theta(z),\end{aligned} \quad (18)$$

where $\Lambda(z)=\int_0^z f_{\lambda_1}(\lambda_1)d\lambda_1=F_{\lambda_1}(z)$, and the terms Θ can be represented as

$$\Theta(z)=\frac{K+1}{\bar{\lambda}_1e^K}e^{-\frac{z}{\bar{\lambda}_2}}\int_0^z e^{-c_1\lambda_1}I_0\left(2\sqrt{\frac{K(K+1)\lambda_1}{\bar{\lambda}_1}}\right)\mathrm{d}\lambda_1, \quad (19)$$

where $c_1=\frac{K+1}{\bar{\lambda}_1}-\frac{1}{\bar{\lambda}_2}>0$. In order to solve (19), we introduce the variable $t^2=c_1\lambda_1$ and substitute $\mathrm{d}\lambda_1=\frac{2t}{c_1}\mathrm{d}t$ into (19). Then, using the specific case of the incomplete Toronto function [14, Eq. (35), Eq. (38)], we can obtain the simplified expression (13) with $c_2=\frac{K(K+1)}{c_1\bar{\lambda}_1}$.

Combining (15) and (18), we can obtain $\mathbb{P}_{\mathrm{FA}}$ and $\mathbb{P}_{\mathrm{MD}}$, which completes the proof of Lemma 1. ■

Therefore, the detection error probability $\mathbb{P}_{\mathrm{E}}$ at $\mathcal{W}$ for a prescribed detection threshold ν can be given by

$$\mathbb{P}_{\mathrm{E}}=\begin{cases}1, & \nu<\sigma^2,\\ 1-\Theta, & \nu\geq\sigma^2,\end{cases} \quad (20)$$

According to $c_1>0$, we have

$$\rho>\frac{\beta_{rw}}{\beta_{rw}+\beta_{sw}(K_{rw}+1)}=\rho_{\min}, \quad (21)$$

which indicates that the power allocation factor has a new range of $\rho\in(\rho_{\min},1)$.

B. Optimal Decision Threshold for $\mathcal{W}$

Since $\mathcal{W}$ aims at minimizing the detection error probability $\mathbb{P}_{\mathrm{E}}=\mathbb{P}_{\mathrm{FA}}+\mathbb{P}_{\mathrm{MD}}$, the case of $\nu<\sigma^2$ will never be chosen because it indicates a completely incorrect detection. Thus, the optimal detection threshold ν^* for $\mathcal{W}$ exists when $\nu\geq\sigma^2$, and can be obtained by solving $\frac{\partial\mathbb{P}_{\mathrm{E}}}{\partial\nu}=0$ in the following lemma.

Lemma 2: The optimal threshold for the detector can be calculated as

$$\frac{1-Q_1(\sqrt{2c_2},\sqrt{2c_1(\nu-\sigma^2)})}{2\bar{\lambda}_2 c_1 e^{c_2+c_1(\nu-\sigma^2)}}-I_0(\sqrt{4c_1c_2(\nu-\sigma^2)})=0. \quad (22)$$

Proof: Taking the derivative of $\mathbb{P}_{\mathrm{E}}$ with respect to ν yields

$$\frac{\partial \mathbb{P}_{\mathrm{E}}}{\partial \nu}=\left[\frac{1-Q_1(\sqrt{2c_2},\sqrt{2c_1(\nu-\sigma^2)})}{\bar{\lambda}_2}+\frac{\partial Q_1(\sqrt{2c_2},\sqrt{2c_1(\nu-\sigma^2)})}{\partial \nu}\right] \times \frac{(K_{rw}+1)e^{c_2}}{c_1\bar{\lambda}_1 e^{K_{rw}}}e^{-\frac{\nu-\sigma^2}{\bar{\lambda}_2}}=0, \quad (23)$$

or equivalently

$$\frac{1-Q_1(\sqrt{2c_2},\sqrt{2c_1(\nu-\sigma^2)})}{\bar{\lambda}_2}+\frac{\partial Q_1(\sqrt{2c_2},\sqrt{2c_1(\nu-\sigma^2)})}{\partial \nu}=0. \quad (24)$$

According to the derivation rules, we have

$$\frac{\partial Q_1(x,y)}{\partial y}=-ye^{-\frac{x^2+y^2}{2}}I_0(xy), \quad (25)$$

where $x=\sqrt{2c_2}$ and $y=\sqrt{2c_1(\nu-\sigma^2)}$. Then, substituting (25) into (24) yields (22). Note that the equation is also mathematically intractable. Thus, the optimal threshold ν^* can be obtained by the bisection search. ■

Under the prescribed optimal detection threshold ν^* obtained from Lemma 2, the minimum detection error probability at $\mathcal{W}$ can be given by

$$\mathbb{P}_{\mathrm{E}}^*=1-\frac{(K_{rw}+1)e^{\frac{\sigma^2-\nu^*}{\bar{\lambda}_2}}}{c_1\bar{\lambda}_1 e^{K_{rw}-c_2}}\left(1-Q_1\left(\sqrt{2c_2},\sqrt{c_1\bar{\lambda}_2(\nu^*-\sigma^2)}\right)\right). \quad (26)$$

Thus, the covertness constraint can be written as $\mathbb{P}_{\mathrm{E}}^* \geq 1-\epsilon_c$. Note that it is extremely difficult to analyze the monotonicity of $\mathbb{P}_{\mathrm{E}}^*$, and we provide some asymptotic cases of $\mathbb{P}_{\mathrm{E}}^*$ in the following remark.

Remark 1: First, if $\rho \to \rho_{\min}$, $c_1 \to 0$ and $c_2 \to \infty$. Thus, the optimal detection threshold ν^* is infinite according to the solution of (22) and the probability of $\mathcal{W}$ to make detection errors approaches 1. Then, as $\rho \to 1$, which is equivalent to the case without jamming, we can derive that $c_1 \to \infty$ and $c_2 \to K_{rw}$, and the minimum detection error probability approaches 0 $\lim_{\rho\to 1}\mathbb{P}_{\mathrm{E}}^*=0$. Moreover, we use the numerical method to observe that $\mathbb{P}_{\mathrm{E}}^*$ first increases, and then monotonically decreases with respect to ρ, which will be confirmed in Fig. 2.

V. Covert Rate Maximization

In this section, we first derive a closed-form expression of the transmission outage probability $\mathbb{P}_{\mathrm{out}}$ for the UAV relaying, based on which we propose power allocation scheme to maximize the covert rate subject to the covertness constraint.

A. Transmission Outage Probability

When $\mathcal{S}$ transmits, the signal received at $\mathcal{D}$ after AF relaying can be given by

$$y_d(i)=g_{rd}\sqrt{\beta_{rd}G}\left[\sqrt{\beta_{sr}P_s}g_{sr}x_s(i)+r(i)+n_r(i)\right]+n_d(i), \quad (27)$$

After some algebraic manipulations, the equivalent instantaneous end-to-end SNR at $\mathcal{D}$ can be written as

$$\gamma_{\mathrm{eq}}=\frac{\gamma_{sr}\gamma_{rd}}{\gamma_{sr}+\gamma_{rd}+1}, \quad (28)$$

where $\gamma_{sr}=P_s\beta_{sr}|g_{sr}|^2/(\sigma^2(1+\xi P_r))$ and $\gamma_{rd}=P_r\beta_{rd}|g_{rd}|^2/\sigma^2$ represent the instantaneous received SNRs on the $\mathcal{S}$-$\mathcal{R}$ and $\mathcal{R}$-$\mathcal{D}$ links, respectively. In the noise-limited system, the outage probability is defined as the probability that the channel capacity $C_{sd}=\log_2(1+\gamma_{\mathrm{eq}})$ falls below the transmission rate R_{sd}, i.e., $C_{sd}<R_{sd}$. Consequently, the outage probability can be given by

$$\mathbb{P}_{\mathrm{out}}=\Pr\left[\gamma_{\mathrm{eq}}<2^{R_{sd}}-1\right]=\Pr\left[\gamma_{\mathrm{eq}}<\bar{\gamma}_{sd}\right]=F_{\gamma_{\mathrm{eq}}}(\bar{\gamma}_{sd}), \quad (29)$$

where $F_{\gamma_{\mathrm{eq}}}(\cdot)$ is the CDF of γ_{eq}. However, the analytical evaluation of the outage probability using γ_{eq} is complicated. To facilitate the derivation of the SNR statistics, γ_{eq} can be approximated by its upper bound γ_{ub} as [15]

$$\gamma_{\mathrm{eq}} \leq \gamma_{ub}=\min(\gamma_{sr},\gamma_{rd}). \quad (30)$$

Thus, a closed-form lower bound to (29) can be given by

$$\begin{aligned}\mathbb{P}_{\mathrm{out}}&=1-\Pr(\gamma_{sr}>\bar{\gamma}_{sd})\Pr(\gamma_{rd}>\bar{\gamma}_{sd})\\&=1-Q_1\left(\sqrt{2K_{sr}},\sqrt{2(K_{sr}+1)(1+\xi P_r)\bar{\gamma}_{sd}\sigma^2/(P_s\beta_{sr})}\right)\\&\quad\times Q_1\left(\sqrt{2K_{rd}},\sqrt{2(K_{rd}+1)\bar{\gamma}_{sd}\sigma^2/(P_r\beta_{rd})}\right).\end{aligned} \quad (31)$$

The outage probability monotonically decreases with the increasing P_s and P_r, and the total transmit power is fixed. This indicated that an optimal power allocation factor ρ_{OP} can be calculated to minimize the outage probability by using the following lemma

Lemma 3: The closed-form solution ρ_{OP} for minimizing outage probability can be written as

$$\left(\frac{\gamma_a}{\gamma_b}\right)^{\frac{3}{4}}\frac{K_{rd}\Xi^{\frac{1}{4}}}{K_{sr}}\left(\frac{P_r}{P_s}\right)^{\frac{7}{4}}\exp\left(\sqrt{\frac{4\gamma_a}{P_s\Xi}}-\sqrt{\frac{4\gamma_b}{P_r}}+K_{rd}-K_{sr}\right)=1, \quad (32)$$

where $\gamma_a=K_{sr}(K_{sr}+1)\bar{\gamma}_{sd}\sigma^2(1+\xi P_t)/\beta sr$, $\gamma_b=K_{rd}(K_{rd}+1)\bar{\gamma}_{sd}\sigma^2/\beta_{rd}$ and $\Xi=(1+\xi P_t)/(1+\xi P_r)$.

Proof: Refer to the proof of Theorem 1 in [16]. ■

B. Solution to the Optimization Problem

Note that the objective function of the optimization problem (7) is equivalent to minimizing the outage probability. After obtaining the power allocation factor ρ_{OP} according to Lemma 3, we need to further discuss the relationship between ρ_{OP} and the feasible region generated by the covertness constraint $\mathbb{P}_{\mathrm{E}}^* \geq 1-\epsilon_c$, and then derive the global optimum solution ρ^*.

According to the analysis in Remark 1, for a given ϵ_c, the covertness constraint can derive a feasible region because the limiting value of $\mathbb{P}_{\mathrm{E}}^*$ is different at the two ends of $(\rho_{\min},1)$, i.e., $\lim_{\rho\to 1}\mathbb{P}_{\mathrm{E}}^* < \lim_{\rho\to\rho_{\min}}\mathbb{P}_{\mathrm{E}}^*$. In other words, the covertness constraint can yield one feasible region of ρ to ensure the covertness, i.e., $\mathcal{A}=\{(\rho_{\min},\rho_C]\}$, where ρ_C denotes the solution to the equation $\mathbb{P}_{\mathrm{E}}^*=1-\epsilon_c$. ρ_C corresponds to the root when the covertness constraint takes the equation, and the root can be calculated by using the bisection method. If $\rho_{OP}\in\mathcal{A}$, the optimal factor is $\rho^*=\rho_{OP}$. Otherwise, the optimal factor is $\rho^*=\rho_C$.

VI. Numerical Results and Discussion

In this section, we present simulation results to evaluate the effectiveness of the proposed scheme under covertness constraints. Without loss of generality, we assume that $\mathcal{S}$, $\mathcal{R}$,

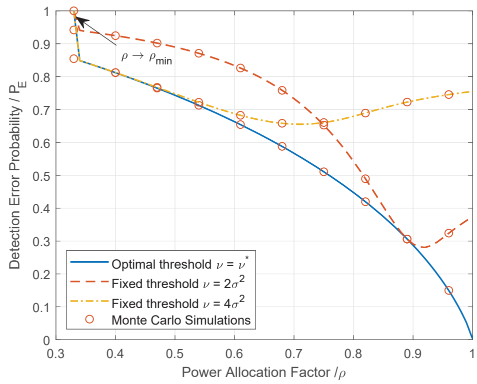

Fig. 2. Detection error probability $\mathbb{P}_{\mathrm{E}}$ versus the power allocation factor ρ under different detection thresholds of $\nu = \nu^{*}$, $\nu = 2\sigma^2$ and $\nu = 4\sigma^2$.

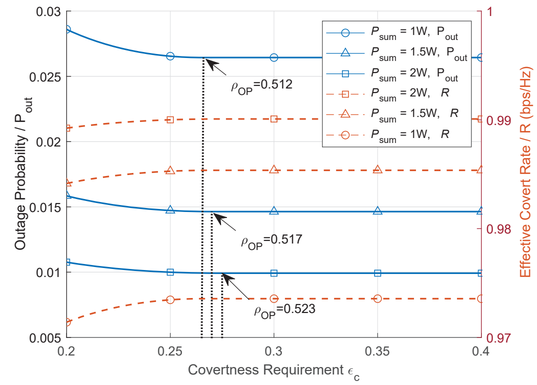

Fig. 3. Outage probability $\mathbb{P}_{\mathrm{out}}$ and effective covert rate R versus ϵ_c for different P_{sum}.

$\mathcal{W}$ and $\mathcal{D}$ are located at $[-200, 0, 0]$, $[0, 0, 50]$, $[0, -300, 0]$ and $[200, 0, 0]$ in meters, respectively. The total transmit power $P_{\mathrm{sum}} = 1$ W and the required transmission rate $R_{sd} = 1$ bit/s/Hz. The channel power gain at the unit distance is $\beta_0 = -60$ dB, and the noise power is $\sigma^2 = -110$ dBm. Other parameters related to the channel and environment are set to $a = 11.95, b = 0.14, \xi_1 = 1, \xi_2 = 2, \kappa_0 = 5$ dB, $\kappa_{\frac{\pi}{2}} = 15$ dB, and $\xi = 0.1$.

Fig. 2 plots the detection error probability $\mathbb{P}_{\mathrm{E}}$ versus ρ for different detection thresholds. It is observed that the minimum detection error probability $\mathbb{P}_{\mathrm{E}}^{*}$ can be obtained when we have the optimal $\nu = \nu^{*}$. In addition, the fixed thresholds of $\nu = 2\sigma^2$ and $\nu = 4\sigma^2$ perform well when ρ is taken large and small, respectively. The Monte-Carlo results match well with the theoretical values according to (14) and (15), respectively. Furthermore, we find that an increase in P_j can create more randomness, which produces a larger detection error probability at $\mathcal{W}$. We also note that when $\rho \to 1$, $\mathbb{P}_{\mathrm{E}}^{*}$ approaches 0, which verifies Remark 1.

Then, we show the performance of the proposed scheme in Fig. 3, which reflects the impact of covertness requirement on the objective function. It can be observed that as ϵ_c increases, R first increases and then remains unchanged. The outage probability varies in the opposite trend to the effective covert rate. When ϵ_c is relative small, the ρ_C calculated by the covertness constraint increases with ϵ_c, and the maximum value of R can be directly obtained by ρ_C, leading to the increasing of R. As ϵ_c continues to increase, the maximum value of R can be obtained by ρ_{OP}, leading to an unchanged R. One can also see from Fig. 3 that as P_{sum} increases from 1 W to 2 W, the effective covert rate improves significantly.

VII. Conclusion

In this work, we have investigated the power optimization of covert communication with the aid of a UAV relay. A cooperative relaying scheme is proposed where the full-duplex UAV relay is adopted for receiving and jamming. The goal is to maximize the effective convert throughput with the covertness constraint. We first find the optimal detection threshold of the detector and calculate the minimum detection error probability. Then, a closed-form expression of outage probability via UAV relaying is derived over Rician fading. Simulation results show that the optimal tradeoff between the information power and jamming power can be found to achieve the network covertness effectively.

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