

Covert Communication in Integrated Satellite Aerial Terrestrial Networks: A NOMA-Enhanced Framework With Practical Constraints

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Abstract—Integrated satellite aerial terrestrial networks (ISATNs) are characterized by their wide-area coverage, which, while enabling massive connectivity, also renders the network more vulnerable to eavesdropping attacks. This inherent security threat necessitates the design of covert communication (CC) schemes to safeguard transmission privacy. Meanwhile, the limited spectrum resource severely restricts the number of accessible users. To tackle this challenge, this paper investigates CC enhanced by non-orthogonal multiple access (NOMA) under practical system imperfections, including hardware impairments (HIs) and channel estimation errors (CEEs). Considering the wide coverage of the satellite beams and the presence of multiple aerial nodes, a threshold-based link selection scheme is proposed to balance system performance and complexity. Closed-form and asymptotic expressions for the detection error probability (DEP), covert outage probability (COP), and the covert transmission rate (CTR) are derived and verified through the Monte Carlo (MC) simulations.

Index Terms—Covert communication, integrated satellite aerial terrestrial networks, practical constraints, non-orthogonal multiple access, threshold-based selection scheme, covert performance.

I. INTRODUCTION

NEXT-GENERATION wireless systems aim to provide wider coverage, higher energy efficiency (EE), and enhanced stability [1]. Satellite communication (Satcom) has

attracted attention due to its wide coverage [2], [3], but direct satellite-to-ground links are often unreliable because of obstacles or adverse weather, such as heavy rain or dense clouds [4], [5]. Relay-assisted transmission, particularly using autonomous aerial vehicles (AAVs), can enhance connectivity and channel quality due to their mobility and flexible deployment [6]. By integrating satellite, aerial, and terrestrial networks, integrated satellite aerial terrestrial networks (ISATNs) offer a promising solution [7], [8], supporting large-scale Internet of Things (IoT) applications, emergency communications, and high-mobility scenarios [9], [10]. However, the explosive growth of devices and data traffic poses significant challenges to spectrum efficiency in ISATNs [11], [12]. To address this issue, non-orthogonal multiple access (NOMA) has emerged as a promising multiple access technique that enables multiple users to access common time-frequency resources by exploiting power-domain multiplexing and successive interference cancellation (SIC) [13], [14], which significantly improves spectral efficiency and supports massive connectivity.

In practical ISATNs, channel estimation errors (CEEs) from imperfect acquisition and hardware impairments (HIs) from non-ideal transceiver parts both reduce system performance and limit SIC, causing residual interference [15], [16], [17]. Ignoring these issues can lead to overly optimistic performance analysis, underscoring the need to include them in practical evaluations. Building on these practical considerations, the open nature of ISATNs significantly increases the risk of eavesdropping, making covert communication (CC) essential [18], [19], [20], which aims to conceal the existence of transmissions from a warden [21]. Notably, practical imperfections and residual interference can further affect the wardens detection capability, which can be exploited to enhance system covertness [22].

Thus, investigating CC in the NOMA-aided ISATNs under practical constraints has become the primary focus of this work. Specifically, a growing number of recent studies address this area, with key advancements and emerging trends summarized in the following sections.

A. Related Works

Following this motivation, ISATNs have attracted significant research attention. Existing studies have extensively

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investigated the performance of ISATNs. The authors in [23] investigated autonomous aerial vehicle (AAV)-assisted ISATNs for mobile edge computing (MEC) with joint optimization of offloading and resource allocation. In [24], channel modeling for ISATNs was investigated, and a unified ISATNs-based approach was proposed to characterize complex communication links. In [25], the authors studied the simultaneously transmitting and reflecting reconfigurable intelligent surface (STAR-RIS)-assisted ISATNs with NOMA under imperfect channel state information (CSI) and analyzed the outage probability (OP). In [26], the authors studied the OP and EE in reconfigurable intelligent surface (RIS)-assisted ISATNs with NOMA under co-channel interference.

To further enhance spectral efficiency and support massive connectivity, NOMA has been integrated into ISATNs, which has attracted increasing research interest. In [27], the authors studied the OP in NOMA-based ISATNs by considering practical channel impairments. In [28], the authors proposed joint power allocation in ISATNs to enhance coverage and spectral efficiency. The authors in [29] studied the ergodic capacity (EC) in NOMA-aided cognitive ISATNs under overlay mode. The authors in [30] studied uplink transmission in ISATNs and developed an optimization algorithm to maximize EE. In [31], the authors investigated the OP and a NOMA power allocation scheme in ISATNs. In [32], the authors investigated a NOMA-aided integrated sensing and communication (ISAC) scheme for ISATNs. In [33], the authors examined the effects of NOMA-based ISATNs equipped with an active RIS.

To enhance secure transmission, recent studies integrate CC with NOMA in ISATNs to improve reliability and secrecy under CEEs and HIs. In [34], the authors investigated CC in ISATNs with multiple LEO satellites. In [35], the closed-form expressions of detection error probability (DEP), covert outage probability (COP), and covert transmission rate (CTR) were derived for NOMA-aided ISATNs under imperfect conditions. In [36], the authors investigated CC in ISATNs, deriving closed-form expressions for DEP, OP, and effective covert rate (ECR). The authors in [37] studied CC in ISATNs under practical constraints, including CEEs, HIs, and co-channel interference. The authors in [38] investigated CC in cognitive ISATNs, and closed-form expressions for DEP, COP, and ECR were derived.

B. Motivations

It is worth noting that existing research on CC in NOMA-aided ISATNs under practical constraints remains insufficient. In [39], the OP and throughput of ISATNs with relay selection, HIs, and interference were analyzed. In [40], the OP was derived in NOMA-based cognitive ISATNs with CEEs, HIs, and interference. In [41], the closed-form expression of end-to-end SNR was provided in RIS-aided ISATNs with imperfect CSI. In [42], the authors studied the OP in RIS-aided ISATNs, considering the influence of CEEs, HIs, and co-channel interference. In [43], the authors examined the effects of HIs on ISATNs with the NOMA technique, with system performance further enhanced by deploying RIS. Different from [35], [37], and [38], this work further focuses on reducing

the relay-selection complexity while characterizing the covert performance under practical system imperfections. Based on the existing literature, the combination of low-complexity threshold-based relay selection and covert performance analysis in NOMA-aided ISATNs under HIs and CEEs has not been sufficiently investigated, which motivates the present study.

C. Contributions

Building on prior studies, this work examines the covert performance of NOMA-aided ISATNs under the combined effects of the HIs, CEEs, and co-channel interference. Specifically, a threshold-based link selection scheme is proposed for the considered ISATNs architecture. The key contributions are summarized as

- Firstly, we establish a CC framework in NOMA-aided ISATNs. Due to severe fading and various propagation obstacles, a direct transmission link between the satellite and terrestrial users is unavailable. Besides, practical constraints, including HIs, CEEs, and co-channel interference, are considered. Especially, a threshold-based link selection scheme is proposed to enhance the stability and reliability of the considered system.
- Secondly, to assess the covert performance of the proposed system model with a threshold-based link selection scheme, a closed-form expression for the DEP is derived. Moreover, the optimal detection threshold of the warden is identified and validated through simulation results. The results demonstrate that the antenna count, jammer count, and the level of HIs have significant impacts on the DEP.
- Thirdly, to evaluate the reliability and interference-resistance capability of the proposed system, closed-form expressions for the COP and CTR are derived. Furthermore, in the numerical simulations, the proposed threshold-based link selection scheme is compared with both the random relay link selection scheme (RRLSS) and the optimal relay link selection scheme (ORLSS), highlighting the advantages of our approach. Besides, to characterize the covert performance in a high signal-to-noise ratio (SNR) regime, the asymptotic expressions of the COP are derived, from which the diversity order and array gain of the considered system are further obtained.
- Finally, Monte Carlo (MC) simulations are conducted to validate the accuracy of the analytical derivations and to examine the effects of different system parameters. The results indicate that interference, HIs, and CEEs have significant impacts on the covert performance of the considered system.

The remainder of this paper is structured as follows. In Section II, the system model and problem are established, where the transmission links, CEEs, and the decoding signal-to-interference plus noise and distortion error (SINDER) of each node are characterized. In Section III, a threshold-based link selection scheme is proposed. In Section IV, the covert performance of the considered system is evaluated, and analytical expressions for the DEP, COP, and CTR are obtained. In Section V, the asymptotic analysis is given

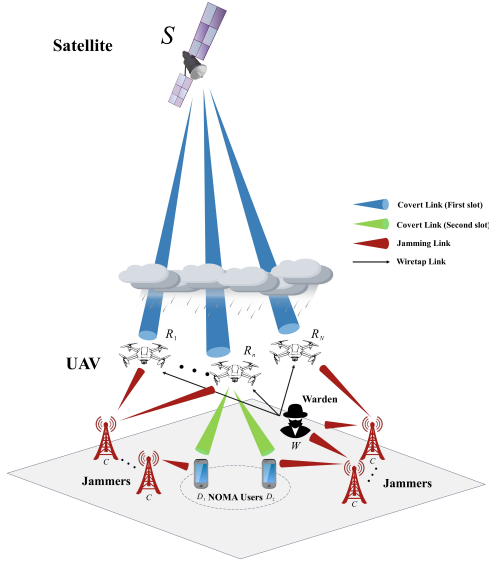


Fig. 1. Description of the system model.

to further illustrate the effects of practical constraints. In Section VI, some numerical results are provided to corroborate the analytical performance evaluation. Finally, in Section VII, a concise summary of the entire work is provided.

Notations: Uppercase bold symbols denote matrices, whereas lowercase bold symbols represent vectors. The Hermitian operation is denoted by $(\cdot)^H$. The Euclidean norm of a vector is expressed as $\|\cdot\|$, and $|\cdot|$ denotes the magnitude of a complex scalar. The symbol $\mathbf{I}_M$ stands for the $M \times M$ identity matrix. The exponential function is written as $\exp(\cdot)$, and $E[\cdot]$ represents the expectation operator. Finally, $\mathbf{C}^{M \times N}$ denotes the set of all $M \times N$ complex positive semidefinite matrices.

II. SYSTEM MODELING AND PROBLEM FORMULATION

In this paper, we establish a reliable CC system in NOMA-aided ISATNs, as depicted in Fig. 1, comprising a multi-beam satellite S , N AAV relays, each equipped with M antennas $R_n, n \in (1 \cdots N)$, a pair of NOMA terrestrial users $D_i, i \in (1, 2)$,¹ a group of cooperative jammers C , and a warden W . Besides, J_1 and J_2 denote the numbers of two distinct groups of cooperative jammers deployed in the first and the second transmission phases, respectively. Specifically, J_1 cooperative jammers generate interference at the AAV relays, while the J_2 cooperative jammers generate interference at the terrestrial users, whose jamming signals can also be received by W . The values of J_1 and J_2 can be the same or different according to the system configuration. Specifically, due to severe blockage caused by buildings or terrain, insufficient direct satellite coverage, or adverse propagation conditions such as heavy rainfall and dense clouds, the direct link between the satellite and the terrestrial users may be severely degraded or unavailable and is therefore treated as non-existent in this work. Accordingly,

¹Without loss of generality, we only consider one pair of NOMA users to facilitate the performance analysis, and the obtained results can be readily extended to multi-pair NOMA scenarios.

the AAVs are deployed as relays to forward signals from the satellite to ground users. For analytical tractability, all AAV relays are assumed to remain stationary, i.e., hovering at fixed locations, during the considered signal transmission. In addition, we adopt the decode-and-forward (DF) relaying protocol in this work. Accordingly, the entire transmission proceeds in two phases, namely the satellite-to-AAV phase and the AAV downlink forwarding phase.

A. The Transmission Between Satellite and Aerial Platform

In the first phase, the transmitted signal $x_s(t)$ satisfying $E[|x_s(t)|^2] = 1$ from S is forwarded to R_n . Thus, at R_n , the obtained signal has the following expression as

$$y_{r_n}(t) \sqrt{P_s} \mathbf{h}_{sr_n}^H \mathbf{v}_1 [x_s(t) + \eta_{sr_n}] + \sum_{j_1=1}^{J_1} \sqrt{P_{j_1 r_n}} \mathbf{h}_{j_1 r_n}^H \mathbf{v}_1 [x_{j_1}(t) + \eta_{j_1}] + \mathbf{v}_1 \mathbf{n}_{r_n}(t), \quad (1)$$

where P_s is denoted as the transmission power of S , $\mathbf{h}_{sr_n}^H$ denotes the channel coefficient vector from S to the n -th R , $\mathbf{v}_1$ is the receive beamforming (BF) vector with $\|\mathbf{v}_1\|^2 = 1$, η_{sr_n} denotes the HIs noise term satisfying $\eta_{sr_n} \sim (0, k_{sr_n}^2)$. $P_{j_1 r_n}$ represents the jamming power contributed by the j_1 -th interference at the n -th R , $\mathbf{h}_{j_1 r_n}^H$ represents the channel coefficient vector from the j_1 -th C to the n -th R , $x_{j_1}(t)$ denotes the transmitted signal which satisfies $E[|x_{j_1}(t)|^2] = 1$, η_{j_1} is the HIs noise term satisfying $\eta_{j_1} \sim (0, k_{j_1 r_n}^2)$, $\mathbf{n}_{r_n}(t)$ represents an $M \times 1$ additive white Gaussian noise (AWGN) vector satisfying $\mathbf{n}_{r_n}(t) \sim \mathcal{CN}(0, \sigma_{r_n}^2 \mathbf{I}_{M \times 1})$, where $\mathbf{I}_{M \times 1}$ is an $M \times 1$ matrix. Besides, owing to the NOMA scheme, the signal sent by S to D_i forwarded by R_n can be expressed by

$$x_s(t) = a_1 x_1(t) + a_2 x_2(t), \quad (2)$$

where a_1 and a_2 denote the coefficients corresponding to the NOMA users D_1 and D_2 , respectively. Besides, it is assumed that the channel condition D_1 is worse than D_2 , which leads to $a_1 > a_2$ along with the constraint $a_1^2 + a_2^2 = 1$. $x_1(t)$ and $x_2(t)$ represent the signal transmitted from S to D_1 and D_2 , respectively, satisfying $E[|x_1(t)|^2] = E[|x_2(t)|^2] = 1$.

B. The Transmission Between Aerial Platform and Terrestrial Nodes

In the second phase, the selected n -th R forwards the decoded signal to the NOMA users D_i based on the DF protocol.² Thus, the received signal at D_i can be expressed by

$$y_{d_i}(t) = \sqrt{P_r} \mathbf{h}_{r_n d_i}^H [\mathbf{v}_2 (b_1 x_1(t) + b_2 x_2(t) + \eta_{r_n d_i})] + \sum_{j_2=1}^{J_2} \sqrt{P_{j_2 d_i}} \mathbf{h}_{j_2 d_i} [x_{j_2 d_i}(t) + \eta_{j_2 d_i}] + n_{d_i}(t), \quad (3)$$

²Since D_1 and D_2 are paired for NOMA transmission, they are considered to be in the same spatial cluster with approximately aligned channel directions. Accordingly, the selected relay employs a common unit-norm maximal-ratio transmitting (MRT) precoding vector $\mathbf{v}_2$ to transmit the superposed signal.

where P_r represents the transmission power of the n -th R , $\mathbf{h}_{r_n d_i}^H$ denotes the channel coefficient vector from the n -th R to the NOMA user D_i , $\mathbf{v}_2$ is the common unit-norm transmit precoding vector for the paired NOMA users, satisfying $\|\mathbf{v}_2\| = 1$. In addition, b_1 and b_2 are also defined as the NOMA power allocation coefficients, satisfying $b_1^2 + b_2^2 = 1$. In accordance with the NOMA principle, $b_1 > b_2$ is set, which is consistent with the relationship between a_1 and a_2 . Moreover, $\eta_{r_n d_i}$ is the HIs noise term following $\eta_{r_n d_i} \sim (0, b_1^2 k_{r_1}^2 + b_2^2 k_{r_2}^2)$, $P_{j_2 d_i}$ represents the jamming power of the j_2 -th interference at D_i , $h_{j_2 d_i}$ denotes the channel coefficient from the j_2 -th interference to D_i . The jamming signal transmitted by C is denoted by $x_{j_2 d_i}(t)$, which satisfies $E[|x_{j_2 d_i}(t)|^2] = 1$. Furthermore, $\eta_{j_2 d_i}$ is the HIs noise term with $\eta_{j_2 d_i} \sim (0, k_j^2)$, and $n_{d_i}(t)$ denotes the AWGN at D_i satisfying $n_{d_i} \sim \mathcal{CN}(0, \sigma_{d_i}^2)$.

C. The Detection Between W and R

Due to the openness of the wireless network environment, any user within the signal coverage of the AAV can receive the signal transmitted from the selected AAV relay R_n . In our system model, W is assumed to be geographically located near the terrestrial NOMA users. Thus, similar to D_i , W can also receive the forwarded signal. Meanwhile, to interfere with the detection judgment of W , a count of J_2 cooperative jammers C are adopted to continuously transmit a jamming signal. Since W always attempts to detect whether the satellite S is secretly communicating with the NOMA users, the signal received by W can be categorized into two cases according to the observation results, which are denoted as H_0 and H_1 . Specifically, H_0 represents the case where S does not transmit covert signals to NOMA users D_i secretly, and thus the signal received at W only consists of the jamming signal sent by C . In contrast, H_1 denotes the hypothesis that S transmits covert signals to D_i secretly, so the signal received at W becomes the superposition of the covert signal from S and the jamming signal from C . Accordingly, under the two hypotheses, the received signal at W can be written as

$$y_w(t) = \begin{cases} \sum_{j_2=1}^{J_2} \sqrt{P_{j_2 w}} h_{j_2 w} [x_{j_2 w}(t) + \eta_{j_2 w}] + n_w(t), & H_0 \\ \sqrt{P_r} \mathbf{h}_{r_n w}^H \mathbf{v}_2 [b_1 x_1(t) + b_2 x_2(t) + \eta_{r_n}] \\ + \sum_{j_2=1}^{J_2} \sqrt{P_{j_2 w}} h_{j_2 w} [x_{j_2 w}(t) + \eta_{j_2 w}] + n_w(t), & H_1, \end{cases} \quad (4)$$

where $P_{j_2 w}$ represents the jamming power of the j_2 -th interference at W , $h_{j_2 w}$ denotes the channel coefficient between the j_2 -th interference and W , $x_{j_2 w}(t)$ is the signal sent by C satisfying $E[|x_{j_2 w}(t)|^2] = 1$, $\eta_{j_2 w}$ and η_{r_n} are the HIs noise terms satisfying $\eta_{j_2 w} \sim \mathcal{CN}(0, k_j^2)$, $\eta_{r_n} \sim \mathcal{CN}(0, b_1^2 k_{r_1}^2 + b_2^2 k_{r_2}^2)$, $n_w(t)$ is the AWGN at W with

³For analytical tractability, all AAV relays are assumed to employ the same transmit power, which simplifies the subsequent derivation of the system performance.

$n_w(t) \sim \mathcal{CN}(0, \sigma_w^2)$. To evaluate the robustness of the considered CC system, a worst-case detection condition is considered for W .⁴

D. The Channel Estimation Errors of the Considered System

In addition to the HIs as a practical non-ideal factor, CEEs are also taken into account. Thus, the channel coefficients between different nodes can be expressed as

$$h_\varphi = \tilde{h}_\varphi + e_{h_\varphi}, \varphi \in \{sr_n, r_n d_i, r_n w\}, \quad (5)$$

where $\tilde{h}_\varphi$ stands for the estimated channel coefficient, while h_φ denotes the practical one. According to [35], $\tilde{h}_\varphi$ and h_φ are assumed to exhibit the same ergodic property, while e_{h_φ} denotes the CEEs orthogonal to $\tilde{h}_\varphi$, which is given by

$$e_{h_\varphi} \sim \mathcal{CN}\left(0, \Phi_{e_{h_\varphi}}\right), \quad \Phi_{e_{h_\varphi}} = \frac{1}{K_\varphi \bar{\Upsilon}_\varphi + 1}, \quad (6)$$

where K_φ denotes the number of estimation symbols, and $\bar{\Upsilon}_\varphi = E[\Upsilon_\varphi] = \frac{P_z E[|h_\varphi|^2]}{N_\Psi}$ with $z \in \{s, r\}$ and $\Psi \in \{S, R_n, D_i, W\}$ represents the average SNR of the estimation symbols. Moreover, $P_z = (1 - \tau) P_{total}$, where $\tau \in (0, 1)$ is the scaling factor. Here, P_{total} stands for the total power, and N_Ψ denotes the AWGN at the corresponding transmitter nodes.

According to [40], $\Phi_{e_{h_\varphi}}$ is expressed as

$$\Phi_{e_{h_\varphi}} = \frac{\bar{\Upsilon}_{id\varphi} k_z^2}{\bar{\Upsilon}_{id\varphi} (k_z^2 + K_\varphi) + 1}, \quad (7)$$

where $\bar{\Upsilon}_{id\varphi}$ is the SNR of the estimation symbol under ideal hardware conditions, and k_z denotes the HIs level.

E. The Decoding SINDER at Each Node

In this section, the decoding SINDER at each node is provided. Since the entire signal transmission process is divided into two phases and the NOMA scheme is adopted in this paper, with the help of (1), the SINDER for decoding $x_1(t)$ at the AAV relay R_n can be expressed as

$$\gamma_{R_n x_1} = \frac{a_1^2 \lambda_{SR_n}}{(a_2^2 + k_{sr_n}^2) \lambda_{SR_n} + (1 + k_{j_1 r_n}^2) \lambda_{J_1 R_n} + E}, \quad (8)$$

where $\lambda_{SR_n} = \frac{P_s \|\tilde{\mathbf{h}}_{sr_n}\|_F^2}{\sigma_{r_n}^2}$, $\lambda_{J_1 R_n} = \frac{\sum_{j_1=1}^{J_1} P_{j_1 r_n} |\mathbf{v}_1 \mathbf{h}_{j_1 r_n}^H|^2}{\sigma_{r_n}^2}$, $E = \frac{P_s \Phi_{e_{h_{sr_n}}} (1 + k_{sr_n}^2)}{\sigma_{r_n}^2} + 1$.

Assuming perfect SIC is applied at R_n , after decoding x_1 , R_n will treat it as noise and cancel it before decoding

⁴Since $|\mathbf{h}_{r_n w}^H \mathbf{v}_2|^2 \leq \|\mathbf{h}_{r_n w}\|_F^2$, in the following DEP analysis, the upper bound $\|\mathbf{h}_{r_n w}\|_F^2$ is adopted as the effective channel gain of W , corresponding to the strongest possible detection capability of W . Therefore, the resulting DEP provides a conservative lower bound on the covertness performance of the considered system.

x_2 . Therefore, the SINDER for decoding x_2 at R_n can be expressed as

$$\gamma_{R_n x_2} = \frac{a_2^2 \lambda_{SR_n}}{k_{sr_n}^2 \lambda_{SR_n} + (1 + k_{j_1 r_n}^2) \lambda_{J_1 R_n} + E}. \quad (9)$$

Since the channel condition of D_1 is assumed to be worse than that of D_2 , signal $x_1(t)$ must be decoded first at D_2 , and then perfect SIC is applied to cancel $x_1(t)$ before decoding $x_2(t)$. With the help of (3), the SINDER for decoding $x_1(t)$ at D_2 can be expressed as

$$\gamma_{D_2 x_1} = \frac{b_1^2 \lambda_{R_n D_2}}{(b_2^2 + A - 1) \lambda_{R_n D_2} + (1 + k_J^2) \lambda_{J_2 D_2} + D}, \quad (10)$$

where $\lambda_{R_n D_2} = \frac{P_r \|\tilde{h}_{r_n d_2}\|_F^2}{\sigma_{d_2}^2}$, $\lambda_{J_2 D_2} = \frac{\sum_{j_2=1}^{J_2} P_{j_2 d_2} |h_{j_2 d_2}|^2}{\sigma_{d_2}^2}$, $D = \frac{AP_r \Phi_{e_{h_{r_n d_2}}}}{\sigma_{d_2}^2} + 1$, $A = 1 + b_1^2 k_{r_1}^2 + b_2^2 k_{r_2}^2$.

Then, the SINDER for decoding $x_2(t)$ at D_2 can be expressed as

$$\gamma_{D_2 x_2} = \frac{b_2^2 \lambda_{R_n D_2}}{(A - 1) \lambda_{R_n D_2} + (1 + k_J^2) \lambda_{J_2 D_2} + D}. \quad (11)$$

Similarly, the SINDER for decoding $x_1(t)$ at D_1 is given by

$$\gamma_{D_1 x_1} = \frac{b_1^2 \lambda_{R_n D_1}}{(b_2^2 + A - 1) \lambda_{R_n D_1} + (1 + k_J^2) \lambda_{J_2 D_1} + C}, \quad (12)$$

where $\lambda_{R_n D_1} = \frac{P_r \|\tilde{h}_{r_n d_1}\|_F^2}{\sigma_{d_1}^2}$, $\lambda_{J_2 D_1} = \frac{\sum_{j_2=1}^{J_2} P_{j_2 d_1} |h_{j_2 d_1}|^2}{\sigma_{d_1}^2}$, $C = \frac{AP_r \Phi_{e_{h_{r_n d_1}}}}{\sigma_{d_1}^2} + 1$.

Besides, with the help of (4), the SINDER for W can be shown as

$$\gamma_W = \begin{cases} (1 + k_J^2) \lambda_{J_2 W} + 1, H_0 \\ A \lambda_{R_n W} + (1 + k_J^2) \lambda_{J_2 W} + B, H_1, \end{cases} \quad (13)$$

where $\lambda_{R_n W} = \frac{P_r \|\tilde{h}_{r_n w}\|_F^2}{\sigma_w^2}$, $\lambda_{J_2 W} = \frac{\sum_{j_2=1}^{J_2} P_{j_2 w} |h_{j_2 w}|^2}{\sigma_w^2}$, $B = \frac{AP_r \Phi_{e_{h_{r_n w}}}}{\sigma_w^2} + 1$.

Accordingly, with the assistance of the DF protocol, the system SINDER for the first phase can be denoted as

$$\gamma_{R_n} = \min [\gamma_{R_n x_1}, \gamma_{R_n x_2}]. \quad (14)$$

Using a similar derivation, the ultimate system SINDER for the second phase can be expressed as

$$\gamma_2 = \min [\min (\gamma_{D_1 x_1}, \gamma_{D_2 x_1}), \gamma_{D_2 x_2}]. \quad (15)$$

Finally, the final SINDER of the considered system is shown as

$$\gamma_{SRD} = \min (\gamma_R, \gamma_2). \quad (16)$$

III. LOW COMPLEXITY SELECTION SCHEME

As mentioned before, the entire transmission process consists of two phases, corresponding to the S - R phase and the R - D phase. The proposed threshold-based link selection scheme is executed in the first phase, and its detailed procedure is described as follows

- Firstly, the satellite broadcasts signals to all relays. The terrestrial scheduling center sets the threshold γ_{req} for relay selection.
- Secondly, the AAV relays are sequentially requested to estimate and feed back their instantaneous SINDER. The terrestrial scheduling center then checks whether the SINDER of the currently examined AAV satisfies $\gamma_{R_n} \geq \gamma_{req}$. Once the first AAV satisfying the threshold is found, it is chosen as the forwarding relay, and the scheduling center sends an instruction to this AAV for signal forwarding. Then, the relay selection procedure terminates.
- Thirdly, if no AAV satisfies the threshold after all candidate AAVs have been examined sequentially, the AAV with the maximum link SINDER is chosen as the forwarding relay as a fallback solution. Similarly, the terrestrial scheduling center sends an instruction to this AAV to perform signal forwarding.
- Finally, the selected AAV relay forwards the received signals to the NOMA users D_1 and D_2 to complete the two-hop cooperative transmission.

Compared with other relay selection schemes, such as the optimal relay selection scheme, the advantages of the proposed threshold-based link selection scheme can be summarized in the following three aspects.

- Firstly, unlike the optimal relay selection, which requires the SINDER information for all AAVs, the proposed scheme stops once a AAV meets the SINDER requirement, thereby reducing unnecessary feedback, delay, and energy consumption, making it better suited for GEO satellite scenarios with long delays and limited link bandwidth. Specifically, the proposed scheme requires between 1 and N relay probes and feedback messages depending on when a qualified relay is found, whereas ORLSS requires the SINDER information of all N candidate relays. In the worst case, where no relay satisfies the threshold, the feedback overhead of the proposed scheme becomes comparable to that of ORLSS.
- Secondly, unlike the optimal scheme that requires global comparison for all N AAVs, the proposed scheme uses sequential threshold-based detection, greatly reducing computational complexity and improving the real-time performance of AAV relay scheduling.
- Thirdly, the proposed scheme enhances AAVs' EE by allowing non-selected AAVs to enter sleep mode, avoiding the continuous sensing and feedback required in optimal relay scheduling, which is ideal for energy-constrained nodes.

Thus, the proposed threshold-based link selection scheme is introduced, with the detailed procedure illustrated in Algorithm 1.

Algorithm 1 Low Complexity Selection Scheme

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1: Input: Number of AAV relays  $N$  and relay selection
   threshold  $\gamma_{req}$ ;
2: Output: SINDER corresponding to the selected link  $\gamma_R$ ;
3: Initialize: Set relay index  $n = 1$ , judgement = false;
4: for  $n = 1$   $N$  do
5:   Request  $R_n$  to estimate and feed back its first-hop
   SINDER  $\gamma_{R_n}$ ;
6:   if  $\gamma_{R_n} \geq \gamma_{req}$  then
7:     Select  $R_n$  as the forwarding relay and set  $\gamma_R =$ 
      $\gamma_{R_n}$ ;
8:     judgement = true; break;
9:   end if
10: end for
11: if judgement == false then
12:   Select  $\gamma_R = \max\{\gamma_{R_1}, \gamma_{R_2}, \dots, \gamma_{R_N}\}$ ;
13:   Set the corresponding relay as the forwarding relay;
14: end if

```

IV. COVERT PERFORMANCE OF CONSIDERED SYSTEM

The covert performance of the considered system is analyzed in this section. Before analyzing the detection performance of W , the modeling expressions for satellite, aerial, and terrestrial communication links are presented. In addition, the probability density function (PDF) and cumulative distribution function (CDF) for each link are derived, respectively.⁵

A. The Satellite Transmission Links

In the considered scenario, we investigate a geosynchronous Earth orbit (GEO) satellite as our evaluation scenario. To efficiently utilize the limited communication resources, we adopt time division multiple access (TDMA), in which the satellite transmission is divided into orthogonal time slots. The channel coefficients $\mathbf{h}_{sr_n}$, which follow the shadowed Rician (SR) fading model, characterize the propagation links from the satellite beam to the AAVs, and are expressed as

$$\tilde{\mathbf{h}}_{sr_n} = C_{sr_n} p_{sr_n}, \quad (17)$$

where C_{sr_n} denotes the random SR component of the satellite-to-AAV channel, and p_{sr_n} is determined by free space loss (FSL) as well as the antenna pattern, which can be written as

$$C_{sr_n} = \sqrt{\frac{H_s H_e f^2}{16\pi^2(l^2 + l_0^2)}}, \quad (18)$$

where f is represented as the carrier wavelength, l represents the distance from satellite S to relay R_n , with $l_0 \approx 3.6 \times 10^4$ Km. Moreover, H_s stands for the satellite beam gain, while H_e refers to the receive antenna gain.

According to [44], the received antenna gain H_e is represented as

$$H_e \text{ (dB)} \simeq \begin{cases} \bar{H}_{\max}, & 0^\circ < \alpha < 1^\circ \\ 32 - 25 \log \alpha, & 1^\circ < \alpha < 48^\circ \\ -10, & 48^\circ < \alpha \leq 180^\circ, \end{cases} \quad (19)$$

⁵Moreover, a quasi-static block fading model is adopted, where the channel coefficients are assumed to remain unchanged within each transmission block.

where $\bar{H}_{\max}$ denotes the maximal beam gain with α as the angle, from [44], H_s has the following expression as

$$H_s \simeq H_s^{\max} \left[36 \frac{J_3(\varrho)}{\varrho^3} + \frac{J_1(\varrho)}{2\varrho} \right]^2. \quad (20)$$

In this formulation, $H_s^{\max}$ represents the maximum beam gain, where $\varrho = 2.07123 \sin \psi / \sin \psi_{3\text{dB}}$, with ψ and $\psi_{3\text{dB}}$ standing for the angles between the corresponding receiver position and the beam center with respect to the satellite and the 3dB angle, respectively. The functions $J_1(\cdot)$ and $J_3(\cdot)$ represent the first-order and third-order Bessel functions of the first kind. Furthermore, the limit $\psi \rightarrow 0$ is considered, yielding $H_s \approx H_s^{\max}$. Accordingly, we set $\mathbf{h}_{sr_n} = C_{sr_n}^{\max} p_{sr_n}$ with $C_{sr_n}^{\max} = \sqrt{\frac{f^2 H_s^{\max} H_e}{16\pi^2(l_0^2 + l^2)}}$.

Then, by utilizing (17) and (19), the PDF of the channel coefficient for a single $S-R$ link can be represented as

$$f(x) = \alpha_0 \sum_{n=0}^{m_0-1} \frac{(-\delta_0)^n (1 - m_{sr})_n}{(n!)^2 (\bar{\lambda}_0)^{n+1}} x^n e^{-\Delta_0 x}, \quad (21)$$

where $\bar{\lambda}_0$ represents the average SNR of the satellite source link. The parameter Δ_0 is defined as $\Delta_0 = \frac{\beta_0 - \delta_0}{\bar{\lambda}_0}$, and $\alpha_0 = \beta_0 \left(\frac{2b_0 m_0}{2b_0 m_0 + \Omega_0} \right)^{m_0}$. Here, β_0 and δ_0 are formulated as $\beta_0 = \frac{1}{2b_0}$ and $\delta_0 = \frac{\beta_0 \Omega_0}{2b_0 m_0 + \Omega_0}$, respectively. Besides, $m_0 > 0$ is defined as the fading severity factor, while $2b_0$ stands for the average power of various factors, and Ω_0 is specified as the average power of the line of sight (LoS) component. In this paper, m_0 is assumed to be an integer value. Furthermore, $(\cdot)_x$ is referred to as the Pochhammer symbol [45].

With the help of [46] and some mathematical steps, the CDF of the channel coefficient indicating the single $S-R$ link can be expressed as

$$F(x) = 1 - \alpha_0 \sum_{n=0}^{m_0-1} \sum_{t=0}^n \frac{(-\delta_0)^n (1 - m_0)_n x^t \exp(-\Delta_0 x)}{n! t! (\bar{\lambda}_0)^{n+1} \Delta_0^{n-t+1}}. \quad (22)$$

Regarding the PDF of λ_{SR_n} in (8), assuming identical channel parameters for all $S-R$ links, it can be expressed as

$$f_{\lambda_{SR_n}}(x) = \sum_{k_1=0}^{m_{SR_n}-1} \cdots \sum_{k_M=0}^{m_{SR_n}-1} \Xi(M) x^{\Lambda_{SR_n}-1} e^{-\Delta_{SR_n} x}, \quad (23)$$

where

$$\Xi(M) \triangleq \prod_{\chi=1}^M \nu(k_\chi) \alpha_{SR_n}^M \prod_{\tau=1}^{M-1} B\left(\sum_{\ell=1}^{\tau} k_\ell + \tau, k_{\tau+1} + 1\right), \quad (24)$$

and $\Lambda_{SR_n} \triangleq \sum_{\chi=1}^M k_\chi + M$, $\nu(k_\chi) = \frac{(1 - m_{SR_n})_{k_\chi} (-\delta_{SR_n})^{k_\chi}}{(k_\chi!)^2 \bar{\lambda}_{SR_n}^{k_\chi+1}}$, $\Delta_{SR_n} = \frac{\beta_{SR_n} - \delta_{SR_n}}{\lambda_{SR_n}}$, $\beta_{SR_n} = \frac{1}{2b_{SR_n}}$, $\alpha_{SR_n} = \beta_{SR_n} \left(\frac{2b_{SR_n} m_{SR_n}}{2b_{SR_n} m_{SR_n} + \Omega_{SR_n}} \right)^{m_{SR_n}}$, $\delta_{SR_n} = \frac{\Omega_{SR_n}}{2b_{SR_n} (2b_{SR_n} m_{SR_n} + \Omega_{SR_n})}$, and $B(\cdot, \cdot)$ depicts the Beta function [45].

With the help of [47], the CDF of λ_{SR_n} has the expression as

$$F_{\lambda_{SR_n}}(x) = 1 - \sum_{k_1=0}^{m_{SR_n}-1} \cdots \sum_{k_M=0}^{m_{SR_n}-1} \sum_{p=0}^{\Lambda_{SR_n}-1} \frac{\Xi(M) (\Lambda_{SR_n} - 1)! x^p}{p! \Delta_{SR_n}^{\Lambda_{SR_n} - p}} e^{-\Delta_{SR_n} x}. \quad (25)$$

B. The Terrestrial Transmission Links

In the considered system, all terrestrial transmission links, including the channels between R_n , W , D_i , and C , are assumed to follow Rayleigh fading. Moreover, by employing MRT and maximal ratio combining (MRC) schemes at the AAVs, the PDFs of $\lambda_{J_1 R_n}$, $\lambda_{J_2 W}$, $\lambda_{J_2 D_i}$, and $\lambda_{R_n D_i}$ can be formulated in a unified expression as

$$f_{\lambda_e}(x) = \sum_{I=1}^{\rho(A_e)} \sum_{J=1}^{\xi_I(A_e)} \frac{\Omega_{I,J}(A_e)}{(J-1)! \bar{\lambda}_{(I)}} x^{J-1} e^{-x/\bar{\lambda}_{(I)}}, \quad (26)$$

where $e \in \{J_1 R_n, J_2 W, J_2 D_i, R_n D_i\}$, $A_e = \text{diag}(\bar{\lambda}_{(1)}, \dots, \bar{\lambda}_{(2)}, \dots, \bar{\lambda}_{(O)})$, $O \in \{M, J_1, J_2\}$, $\rho(A_e)$ denotes the number of distinct diagonal elements of A_e . $\bar{\lambda}_{(1)} > \bar{\lambda}_{(2)}, \dots > \bar{\lambda}_{(\rho(A_e))}$ represents the distinct diagonal elements arranged in decreasing order. Furthermore, $\xi_I(A_e)$ indicates the multiplicity of $\bar{\lambda}_{(I)}$, and $\Omega_{I,J}(A_e)$ denotes the (I, J) -th characteristic coefficient of A_e . Subsequently, with the help of (26) and [48], the CDF for λ_e can be written as

$$F_{\lambda_e}(x) = 1 - \sum_{I=1}^{\rho(A_e)} \sum_{J=1}^{\xi_I(A_e)} \sum_{Y=0}^{J-1} \frac{\Omega_{I,J}(A_e)}{Y!} \left(\frac{x}{\bar{\lambda}_{(I)}} \right)^Y e^{-x/\bar{\lambda}_{(I)}}. \quad (27)$$

C. Detection Error Probability of W

The detection performance of W is examined in this section. Since W is located near the NOMA users, it can also receive the beam transmitted from R_n , thereby further determining whether the satellite has secretly sent covert messages to D_i with the aid of R_n . However, due to the existence of cooperative jammers, even if S does not send covert messages to D_i via R_n , W will still receive interfering signals, which further interferes with the judgment of W . Therefore, the signals received by W can be classified into two cases, denoted as E_1 and E_0 , respectively. Specifically, case E_1 corresponds to the scenario where the received signal energy at W is relatively strong, implying that CC via R_n has been established between S and D_i . In this case, the signals received by W contain both the covert signals from S and the interfering signals. Accordingly, E_0 represents the case where no covert transmission occurs, and W only receives interfering signals. Through long-term observation and by setting a decision threshold,⁶ W can distinguish between the two cases, which gives rise to two detection error probabilities,

⁶With an infinite number of observation samples, the average received energy at W converges to its statistical expectation. By normalizing the received energy and the corresponding detection threshold with respect to the noise power σ_W^2 , the normalized detection statistic γ_W and detection threshold τ_0 are obtained, respectively.

namely the false alarm probability (PFA) and the missed detection probability (PMD). The PFA is defined as the event that W incorrectly judges that covert transmission between the satellite and the NOMA users has occurred, while in fact it hasn't. Mathematically, it can be written in the form

$$P_{FA} = \Pr(\gamma_W > \tau_0 | H_0). \quad (28)$$

In contrast, the PMD is defined as the event that W fails to detect the occurrence of covert transmission between the satellite and the NOMA users, which can be mathematically formulated as

$$P_{MD} = \Pr(\gamma_W < \tau_0 | H_1), \quad (29)$$

where τ_0 represents the decision threshold of W . Then, the DEP of W can be expressed as

$$\xi = P_{FA} + P_{MD}. \quad (30)$$

Theorem 1: The DEP achieved by W can be derived and expressed as

$$\xi = \begin{cases} 1, & \tau_0 < 1 \\ \sum_{u=0}^{J_2-1} \frac{1}{u!} \left(\frac{\tau_0 - 1}{Q} \right)^u \exp\left(-\frac{\tau_0 - 1}{Q}\right), & 1 \leq \tau_0 \leq B \\ \sum_{u=0}^{J_2-1} \frac{1}{u!} \left[\frac{\tau_0 - 1}{Q} \right]^u \exp\left(-\frac{\tau_0 - 1}{Q}\right) \\ + \frac{1}{(J_2 - 1)!} \gamma\left(J_2, \frac{\tau_0 - B}{Q}\right) - \frac{\exp\left(-\frac{\tau_0 - B}{A\bar{\lambda}_{R_n W}}\right)}{(J_2 - 1)! \bar{\lambda}_{J_2 W}^{J_2}} \\ \times \sum_{l=0}^{M-1} \sum_{i=0}^l \frac{1}{l!} \left(-\frac{1 + k_J^2}{A\bar{\lambda}_{R_n W}} \right)^l \\ \times \binom{l}{i} \left(-\frac{\tau_0 - B}{1 + k_J^2} \right)^{l-i} \left[\frac{A\bar{\lambda}_{R_n W} - Q}{A\bar{\lambda}_{R_n W} \bar{\lambda}_{J_2 W}} \right]^{-J_2-i} \\ \times \gamma\left(J_2 + i, \frac{[A\bar{\lambda}_{R_n W} - Q](\tau_0 - B)}{A\bar{\lambda}_{R_n W} Q}\right), & \tau_0 > B. \end{cases}$$

where $Q = \bar{\lambda}_{J_2 W} (1 + k_J^2)$, $\gamma(a, x)$ is the incomplete gamma function defined in [45]. Furthermore, the optimal detection threshold of W , denoted by τ_0^* , lies in the region $\tau_0 > B$ and can be obtained as

$$\tau_0^* = \arg \min_{\tau_0 > B} \xi(\tau_0). \quad (31)$$

Then, the corresponding minimum DEP is given by

$$\xi^* = \xi(\tau_0^*). \quad (32)$$

Proof: A detailed proof of the above Theorem is provided in Appendix A. $\square$

D. Covert Outage Probability of the Considered System

The COP is examined herein as another key metric for evaluating the reliability of the considered system. In the proposed model, the COP is defined as the probability that the system SINDER falls below a predefined threshold γ_0 . If the SINDER of any link is lower than γ_0 , an outage occurs.

Theorem 2: Under the proposed threshold-based link selection scheme, the COP can be formulated as

$$\begin{aligned}
 P_{out}(\gamma_0) &= \Pr(\gamma_{SRD} \leq \gamma_0) = \Pr[\min(\gamma_R, \gamma_2) \leq \gamma_0] \\
 &= \underbrace{\Pr[\gamma_R \leq \gamma_0]}_{Z_1} + \underbrace{\Pr[\gamma_2 \leq \gamma_0]}_{Z_2} - \underbrace{\Pr[\gamma_R \leq \gamma_0]}_{Z_1} \underbrace{\Pr[\gamma_2 \leq \gamma_0]}_{Z_2}. \quad (33)
 \end{aligned}$$

Proof: The detailed steps can be seen in Appendix B. $\square$

E.: Covert Transmission Rate of the Considered System

The CTR performance is evaluated in this section. According to [49], CTR is defined as the optimal information transmission rate that ensures reliable communication between the satellite S and the NOMA user D_i . Therefore, it also serves as an important metric for assessing CC system performance. It is worth noting that interference signals from jamming nodes C can affect the covert transmission between S and D_i .

To characterize CTR, we consider both the channel fading and the interference from cooperative jammers. The achievable rate is determined by the minimum of the SINDER across the relevant links while ensuring that the CC remains undetectable to the adversary. Accordingly, CTR can be formulated as a function of the system parameters, including the transmit powers, channel gains, and the interference level from C .

A detailed analysis of CTR under different system configurations is provided in the following section, which highlights the impact of key parameters such as the number of AAV antennas, power allocation coefficients, and the fading severity.

Theorem 3: The CTR of the considered system is expressed as in (34), shown at the bottom of the page. Besides, x_i represents the i -th zero of Legendre Polynomials, ω_i depicts the weight factors from Table (25.4) in [50], and $\gamma^{\max_1} = \frac{b_1^2}{b_2^2 + A - 1}$, $\gamma^{\max_2} = \frac{b_2^2}{A - 1}$.

Proof: The detailed steps can be seen in Appendix C. $\square$

V. ANALYSIS IN THE ASYMPTOTIC REGIME

To better illustrate the effects of HIs, CEEs, and interference on the considered ISATNs, we analyze the asymptotic behavior under the proposed threshold-based link selection scheme in the high SNR regime.

Theorem 4: The asymptotic COP under high SNR conditions can be formulated as

$$P_{out}(\gamma_0) = Z_1^\infty + Z_2^\infty - Z_1^\infty Z_2^\infty. \quad (35)$$

After detailed analysis, the diversity order is obtained as $\Theta = M$, and the corresponding array gain θ is expressed as

$$\theta = \begin{cases} (Q_1 + Q_2) + P_1 + P_2 + P_3, & \gamma_0 \geq \gamma_{req} \\ P_1 + P_2 + P_3, & \gamma_0 < \gamma_{req}. \end{cases} \quad (36)$$

Proof: The detailed steps can be seen in Appendix D. $\square$

TABLE I
CHANNEL PARAMETERS

Shadowing Condition	m_{SRn}	b_{SRn}	Ω_{SRn}
Frequent Heavy Shadowing (FHS)	1	0.063	0.0007
Average Shadowing (AS)	5	0.251	0.279
Infrequent Light Shadowing (ILS)	10	0.158	1.29

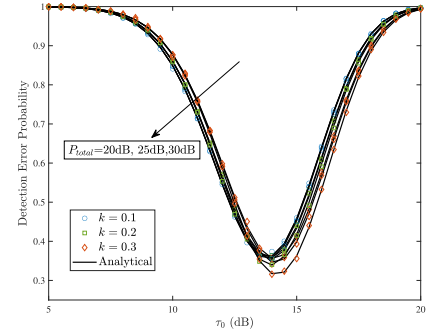


Fig. 2. DEP versus τ_0 under different k and P_{total} with $J = 5$, $M = 2$, $K = 15$, $a_1^2 = 0.7$, $\tau = 0.7$, $\bar{\gamma}_I = 5\text{dB}$, and $\bar{\gamma} = 10\text{dB}$.

VI. NUMERICAL RESULTS

In this section, numerical results are provided to evaluate the influence of key parameters on the performance of reliable CC. The study focuses on how variations in channel conditions, system configurations, and environmental factors impact the ability to ensure secrecy and reliability simultaneously. For clarity, the parameters of the SR fading channels adopted in the simulations are summarized in Table I.

For analytical tractability, we set $k_{srn} = k_{j1rn} = k_{r1} = k_{r2} = k_J = k$, $a_1 = b_1$, $\sigma_{rn}^2 = \sigma_w^2 = \sigma_{di}^2 = 1$, $J_1 = J_2 = J$, $\bar{\lambda}_{SRn} = \bar{\lambda}_{RnDi} = \bar{\lambda}_{RnW} = \bar{\gamma}$, and $\bar{\lambda}_{J1Rn} = \bar{\lambda}_{J2Di} = \bar{\lambda}_{J2W} = \bar{\gamma}_I$. Additionally, the number of MC simulations conducted is set to 10^6 .

Fig. 2 illustrates the DEP versus τ_0 under different k and P_{total} . As shown in the figure, all theoretical solutions are consistent with the simulation results, which verifies the correctness of our theoretical derivation. In terms of the curve trend, the DEP first decreases and then increases as τ_0 increases, and there exists a minimum value. This indicates that the warden has an optimal threshold that minimizes its own DEP. Meanwhile, a smaller k corresponds to a larger DEP, implying that the level of HIs affects the judgment of W on the CC. In addition, P_{total} also reflects the transmit power of the satellite S and the relay R . A higher transmit

⁷The two sets of NOMA power allocation coefficients can generally be configured independently for the two transmission phases. Here, identical values are adopted to reduce the number of varying parameters and facilitate the evaluation of other system parameters.

$$R = \log_2(1 + \gamma^{\max_1}) F_{\gamma_1}(\gamma^{\max_1}) - \frac{\gamma^{\max_1}}{2 \ln 2} \sum_{i=1}^n \frac{\omega_i F_{\gamma_1}(x_i)}{1 + x_i} + \log_2(1 + \gamma^{\max_2}) F_{\gamma_{D2x2}}(\gamma^{\max_2}) - \frac{\gamma^{\max_2}}{2 \ln 2} \sum_{i=1}^n \frac{\omega_i F_{\gamma_{D2x2}}(x_i)}{1 + x_i}. \quad (34)$$

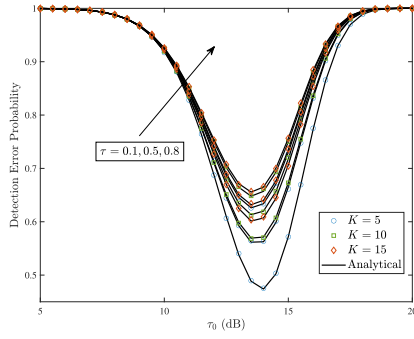


Fig. 3. DEP versus τ_0 under different K and τ with $P_{total} = 20$ dB, $J = 5$, $M = 3$, $k = 0.5$, $a_1^2 = 0.7$, $\bar{\gamma}_I = 5$ dB, and $\bar{\gamma} = 5$ dB.

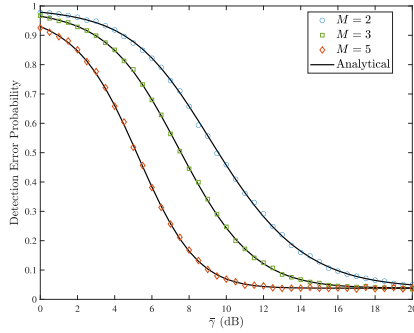


Fig. 4. DEP versus $\bar{\gamma}$ under different M with $\tau_0 = 15$ dB, $P_{total} = 20$ dB, $J = 5$, $\tau = 0.7$, $K = 5$, $k = 0.1$, $a_1^2 = 0.7$, and $\bar{\gamma}_I = 5$ dB.

power leads to a lower DEP, which increases the probability that W can correctly detect the CC.

Fig. 3 depicts the DEP versus τ_0 under different K and τ . As illustrated in the figure, the variation trend of DEP with τ_0 is similar to that in Fig. 2, first decreasing and then increasing with the rise of τ , which also confirms that the warden has an optimal detection threshold. Specifically, a larger K leads to a lower DEP since a greater number of estimation symbols improves the channel estimation accuracy and thus enhances the detection performance of the warden. Meanwhile, a larger τ implies a reduction in the transmit power of the satellite S and the selected relay R , which degrades the received signal quality and results in an increased DEP.

Fig. 4 introduces the DEP versus $\bar{\gamma}$ for different values of M . As shown in the figure, the DEP decreases as $\bar{\gamma}$ increases, indicating that more severe channel fading results in poorer detection performance at W . Moreover, the DEP declines with an increasing number of antennas M . This trend is attributed to the use of MRT at R , which boosts the signal energy received by W as M increases. Consequently, it becomes easier to distinguish between the hypotheses H_0 and H_1 using energy detection, thereby reducing the error probability at W .

Fig. 5 examines the DEP versus $\bar{\gamma}_I$ under different J_2 and τ_0 . It can be seen that the DEP increases with the number of interferences, as greater aggregate interference at W makes detection more difficult. Additionally, as τ_0 increases, the DEP decreases. This trend indicates that different detection thresholds yield different DEP values, implying that W must continuously adjust his decision threshold to achieve

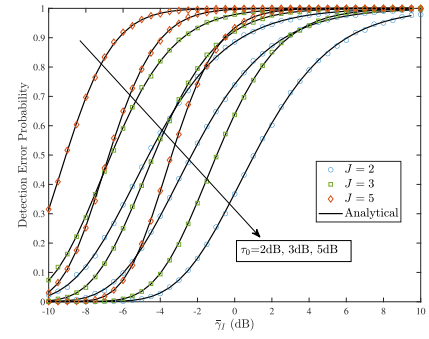


Fig. 5. DEP versus $\bar{\gamma}_I$ under different J_2 and τ_0 with $P_{total} = 20$ dB, $M = 3$, $\tau = 0.7$, $K = 5$, $k = 0.1$, $a_1^2 = 0.7$, and $\bar{\gamma} = 10$ dB.

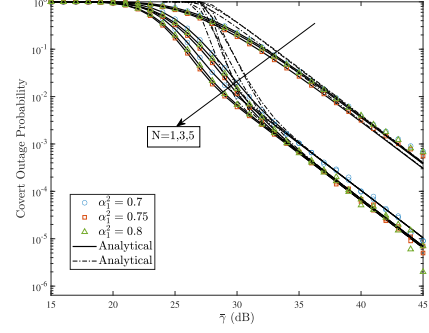


Fig. 6. COP versus $\bar{\gamma}$ under different N and a_1^2 with $P_{total} = 20$ dB, $M = 2$, $\tau = 0.7$, $K = 10$, $k = 0.1$, $\bar{\gamma}_I = 5$ dB, $\gamma_{req} = 3$ dB, and $\gamma_0 = 2$ dB in the FHS scenario.

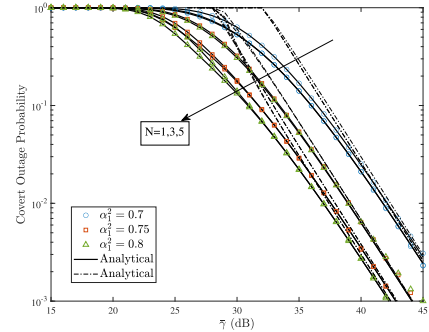


Fig. 7. COP versus $\bar{\gamma}$ under different N and a_1^2 with $P_{total} = 20$ dB, $M = 2$, $\tau = 0.7$, $K = 10$, $k = 0.1$, $\bar{\gamma}_I = 5$ dB, $\gamma_{req} = 2$ dB and $\gamma_0 = 3$ dB in the FHS scenario.

optimal detection performance. Furthermore, the DEP rises with increasing $\bar{\gamma}_I$ and eventually approaches 1, indicating that stronger interference channels significantly impair the wardens detection capability and increase the DEP.

Fig. 6 provides the COP versus $\bar{\gamma}$ under different N and a_1^2 in the FHS scenario. When $\gamma_0 < \gamma_{req}$, the COP is mainly determined by relay selection. It decreases with higher $\bar{\gamma}$ and larger N , which improve the chances of meeting γ_{req} and finding a qualified relay. Additionally, increasing a_1^2 enhances the weaker user D_1 's decoding reliability, further lowering the COP.

Fig. 7 illustrates the COP versus $\bar{\gamma}$ under different N and a_1^2 in the FHS scenario. In this case, $\gamma_0 > \gamma_{req}$, where the performance is dominated by the decoding constraint.

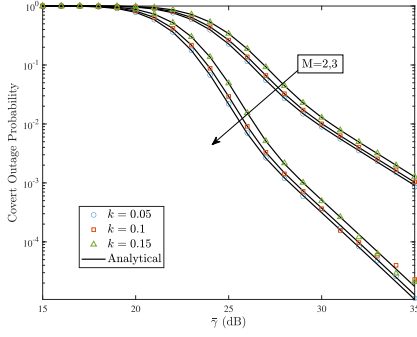


Fig. 8. COP versus $\bar{\gamma}$ under different M and k with $P_{total} = 20\text{dB}$, $N = 5$, $\tau = 0.7$, $K = 10$, $a_1^2 = 0.7$, $\gamma_{req} = 3\text{dB}$ and $\gamma_0 = 2\text{dB}$ in the FHS scenario.

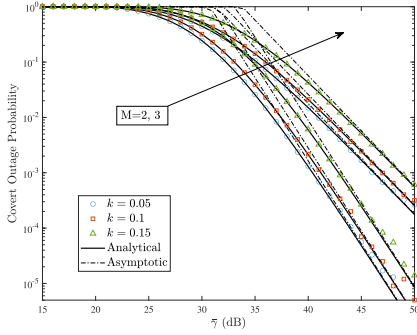


Fig. 9. COP versus $\bar{\gamma}$ under different M and k with $P_{total} = 20\text{dB}$, $N = 5$, $\tau = 0.7$, $K = 10$, $a_1^2 = 0.7$, $\bar{\gamma}_I = 5\text{dB}$, $\gamma_{req} = 2\text{dB}$ and $\gamma_0 = 3\text{dB}$ in the FHS scenario.

Compared with Fig. 6, all curves shift to the right, indicating that a higher SNR is required to satisfy the more stringent threshold. In this regime, even if a AAV satisfies γ_{req} , the decoding SINDER may still fail to meet γ_0 , making the benefit of increasing N less significant. In contrast, increasing a_1^2 remains effective, since allocating more power to the weaker user D_1 directly improves decoding reliability.

Fig. 8 shows the COP versus $\bar{\gamma}$ under different M and k in the FHS scenario. In this scenario, $\gamma_{req} > \gamma_0$, so the performance is dominated by the relay selection constraint. As $\bar{\gamma}$ increases, all curves decline, with the COP approaching zero for $\bar{\gamma} > 30\text{dB}$. Increasing M lowers the COP and shifts the curve leftward due to antenna diversity, which strengthens the received signal and makes it easier to meet the relay threshold γ_{req} . Meanwhile, smaller k values reduce the COP by weakening HIs, mitigating signal distortion and improving link quality.

Fig. 9 depicts the COP versus $\bar{\gamma}$ under different M and k in the FHS scenario. In this scenario, $\gamma_0 > \gamma_{req}$, which means system performance is constrained by the decoding requirement. Compared to Fig. 8, all curves shift to the right, indicating that stronger channel conditions are necessary to satisfy the stricter decoding threshold. Even if a AAV meets the relay selection criterion, decoding failures may still occur, diminishing the advantages of increasing M . However, a larger number of antennas continues to offer diversity gain and lower the COP, while smaller k values remain beneficial by reducing distortion and enhancing decoding reliability.

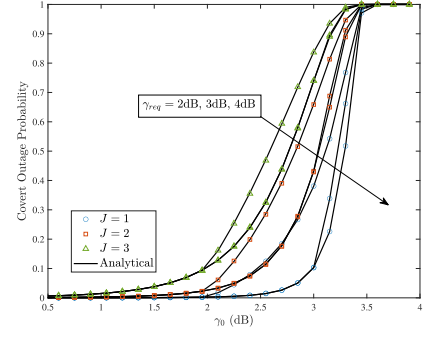


Fig. 10. COP versus γ_0 under different J and γ_{req} with $P_{total} = 20\text{dB}$, $N = 3$, $\tau = 0.7$, $K = 10$, $a_1^2 = 0.7$, and $\bar{\gamma}_I = 5\text{dB}$ in the FHS scenario.

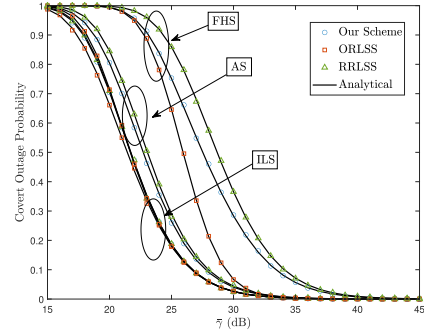


Fig. 11. COP versus $\bar{\gamma}$ under different scenario and scheduling schemes with $P_{total} = 20\text{dB}$, $N = 3$, $J = 2$, $M = 2$, $\bar{\gamma}_I = 5\text{dB}$, $\tau = 0.7$, $k = 0.01$, $K = 10$, $a_1^2 = 0.7$, $\gamma_{req} = 2\text{dB}$ and $\gamma_0 = 3\text{dB}$.

Fig. 10 illustrates the COP versus γ_0 under different J and γ_{req} in the FHS scenario. It is evident that the COP rises from 0 to 1 as γ_0 increases, since a higher outage threshold makes meeting the decoding SINDER requirement harder. More jammers also increase COP due to stronger interference, while a lower γ_{req} eases relay selection, boosting successful transmission and reducing COP.

Fig. 11 examines the COP versus $\bar{\gamma}$ under different scenarios and scheduling schemes. In RRLSS, one relay is uniformly and randomly selected from the N AAV relays without using their instantaneous SINDER information. In ORLSS, the instantaneous first-hop SINDER of all candidate relays is evaluated, and the relay with the maximum SINDER is selected. Moreover, it is worth noting that the proposed scheme consistently outperforms the RRLSS under all three shadowing scenarios and achieves performance close to the ORLSS in both AS and ILS conditions.

Fig. 12 examines the COP versus $\bar{\gamma}_I$ under different k and scheduling schemes. A clear trend can be identified that as $\bar{\gamma}_I$ increases, the COP of all three schemes increases and eventually converges to 1, indicating that stronger interference degrades CC performance. Moreover, the curves show that the proposed scheme consistently outperforms the RRLSS under all values of k , and achieves performance close to the ORLSS when $\bar{\gamma}_I$ is large enough. Furthermore, a larger k leads to a higher COP across all schemes, which implies that increased HIs levels further deteriorate the system's COP.

Fig. 13 plots the CTR versus $\bar{\gamma}$ under different M and k in the FHS scenario. Overall, the CTR rises sharply with $\bar{\gamma}$ and

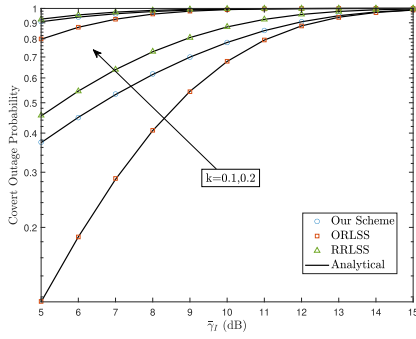


Fig. 12. COP versus $\bar{\gamma}_I$ under different k and scheduling schemes with $P_{total} = 20\text{dB}$, $N = 3$, $J = 3$, $\bar{\gamma} = 30\text{dB}$, $\tau = 0.7$, $K = 10$, $a_1^2 = 0.7$, $\gamma_{req} = 2\text{dB}$ and $\gamma_0 = 3\text{dB}$ in the FHS scenario.

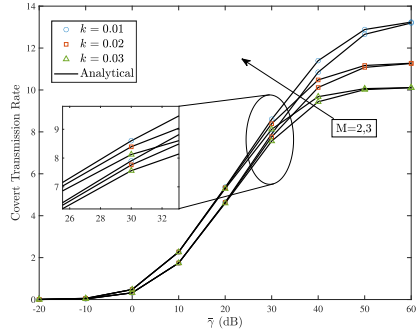


Fig. 13. CTR versus $\bar{\gamma}$ under different M and k with $P_{total} = 20\text{dB}$, $N = 3$, $\tau = 0.7$, $K = 5$, $a_1^2 = 0.7$, $\bar{\gamma}_I = 5\text{dB}$, and $J = 2$ in the FHS scenario.

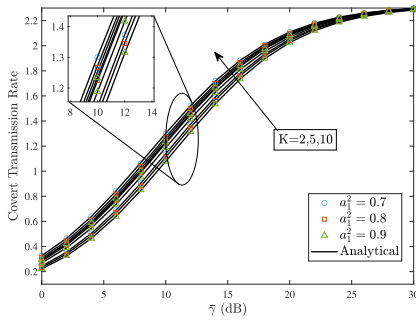


Fig. 14. CTR versus $\bar{\gamma}$ under different a_1^2 and K with $P_{total} = 20\text{dB}$, $M = 3$, $N = 3$, $\tau = 0.7$, $k = 0.5$, $\bar{\gamma}_I = 5\text{dB}$, and $J = 2$ in the FHS scenario.

saturates at high SNR, indicating that higher average SNR mainly drives CTR improvement. Increasing M from 2 to 3 consistently boosts CTR across all SNRs, which improves reception reliability and supports higher data rates. CTR is also highly sensitive to k , with smaller k yielding better performance, highlighting the importance of mitigating HIs.

Fig. 14 illustrates the CTR versus $\bar{\gamma}$ under different a_1^2 and K in the FHS scenario. As can be seen from the figure, similar to Fig. 13, the CTR increases with the average SNR and eventually saturates. In addition, a larger K leads to a higher CTR, as more symbols are used to estimate the channel. Besides, by adjusting the NOMA power allocation factor, a higher CTR can be obtained.

VII. CONCLUSION

This paper has analyzed the covert performance of ISATNs under practical constraints. To improve spectral efficiency, NOMA was incorporated into the transmission framework. Practical deployment conditions, non-ideal factors, including co-channel interference, HIs, and CEEs, are explicitly modeled to reflect real-world imperfect conditions. To improve robustness against interference, a threshold-based link selection scheme was further proposed. Closed-form expressions for the DEP, COP, and CTR were derived, with their accuracy confirmed through MC simulations. Numerical results demonstrate the influence of interference, HIs, and CEEs on the covert performance. However, the proposed scheme still has several limitations. In particular, the relay selection threshold is predetermined, and statistically identical AAV channels are assumed for analytical tractability. Future work will investigate adaptive threshold selection, non-identical AAV channel conditions, and mobile AAV deployment to further improve the adaptability of the proposed framework to more dynamic and heterogeneous ISATNs scenarios. Overall, these findings provide a systematic analytical framework for evaluating CC in ISATNs and offer practical insights for designing robust and spectrally efficient secure transmissions in next-generation integrated networks.

APPENDIX A

PROOF OF THEOREM 1

As mentioned before, the DEP is composed of two components, namely PFA and PMD. Therefore, we sequentially present the derivation processes herein. First, recalling (28), it can be further expressed as

$$\begin{aligned} P_{FA}(\tau_0) &= \Pr[(1 + k_J^2) \lambda_{J_2 W} \geq \tau_0 - 1] \\ &= 1 - \int_0^{\frac{\tau_0 - 1}{1 + k_J^2}} f_{\lambda_{J_2 W}}(x) dx. \end{aligned} \quad (37)$$

Then, by taking (26) into (37), we will get

$$\begin{aligned} P_{FA}(\tau_0) &= \begin{cases} 1, & \tau_0 \leq 1 \\ \sum_{u=0}^{J_2-1} \frac{1}{u!} \left[\frac{\tau_0 - 1}{\bar{\lambda}_{J_2 W} (1 + k_J^2)} \right]^u \exp \left[-\frac{\tau_0 - 1}{\bar{\lambda}_{J_2 W} (1 + k_J^2)} \right], & \tau_0 > 1. \end{cases} \end{aligned} \quad (38)$$

Similarly, recalling (29), it can be expressed as

$$\begin{aligned} P_{MD}(\tau_0) &= \Pr \left[\lambda_{R_n W} \leq \frac{\tau_0 - B - (1 + k_J^2) \lambda_{J_2 W}}{A} \right] \\ &= T_1 - T_2, \end{aligned} \quad (39)$$

where

$$\begin{aligned} T_1 &= \int_0^{\frac{\tau_0 - B}{1 + k_J^2}} f_{\lambda_{J_2 W}}(x) dx, \\ T_2 &= \int_0^{\frac{\tau_0 - B}{1 + k_J^2}} f_{\lambda_{J_2 W}}(x) \sum_{l=0}^{M-1} \frac{1}{l!} \left[\frac{\tau_0 - B - (1 + k_J^2) x}{A \bar{\lambda}_{R_n W}} \right]^l \end{aligned} \quad (40)$$

$$\times \exp \left[-\frac{\tau_0 - B - (1 + k_J^2) x}{A \bar{\lambda}_{R_n} W} \right] dx. \quad (41)$$

Then, by taking (26) into (40) and (41), we will get

$$P_{MD}(\tau_0) = \begin{cases} 0, & \tau_0 \leq B \\ \frac{1}{(J_2 - 1)!} \gamma \left(J_2, \frac{\tau_0 - B}{\bar{\lambda}_{J_2} W (1 + k_J^2)} \right) - \frac{\exp \left(-\frac{\tau_0 - B}{A \bar{\lambda}_{R_n} W} \right)}{(J_2 - 1)! \bar{\lambda}_{J_2} W} \\ \times \sum_{l=0}^{M-1} \sum_{i=0}^l \frac{1}{l!} \left(-\frac{1 + k_J^2}{A \bar{\lambda}_{R_n} W} \right)^l \binom{l}{i} \left(-\frac{\tau_0 - B}{1 + k_J^2} \right)^{l-i} \\ \times \left[\frac{A \bar{\lambda}_{R_n} W - (1 + k_J^2) \bar{\lambda}_{J_2} W}{A \bar{\lambda}_{R_n} W \bar{\lambda}_{J_2} W} \right]^{-J_2-i} \\ \times \gamma \left(J_2 + i, \frac{[A \bar{\lambda}_{R_n} W - (1 + k_J^2) \bar{\lambda}_{J_2} W] (\tau_0 - B)}{A \bar{\lambda}_{R_n} W \bar{\lambda}_{J_2} W (1 + k_J^2)} \right), & \tau_0 > B. \end{cases} \quad (42)$$

Then, by taking (38) and (42) into (30), (31) will be obtained. To further deduce the optimal detection threshold of W , the monotonicity in each range of (31) should be analyzed. For $\tau_0 < 1$, $\frac{\partial \xi}{\partial \tau_0} = 0$. For $1 \leq \tau_0 \leq B$, $\frac{\partial \xi}{\partial \tau_0}$ is given by

$$\frac{\partial \xi}{\partial \tau_0} = -\frac{1}{Q(J_2 - 1)!} \left(\frac{\tau_0 - 1}{Q} \right)^{J_2-1} \exp \left(-\frac{\tau_0 - 1}{Q} \right) < 0. \quad (43)$$

Therefore, ξ monotonically decreases with respect to τ_0 in this region. Moreover, the derivative remains negative at $\tau_0 = B$, indicating that the minimum DEP cannot be achieved within $\tau_0 \leq B$. Hence, the optimal detection threshold satisfies $\tau_0^* > B$, which can be obtained by numerical search, and can be expressed by $\tau_0^* = \arg \min_{\tau_0 > B} \xi(\tau_0)$. Then, by taking τ_0^* into (31), the corresponding minimum DEP will be obtained.

APPENDIX B PROOF OF THEOREM 2

When calculating COP, with the help of (14), (15), and (16), it is important to obtain Z_1 and Z_2 , respectively. Firstly, we give the expression of Z_1 .

$$Z_1 = \begin{cases} \Pr \left[\max \{ \gamma_{R_1}, \gamma_{R_2}, \dots, \gamma_{R_N} \} < \gamma_{req} \right] \\ + \sum_{n=2}^N \Pr \left[\max \{ \gamma_{R_1}, \gamma_{R_2}, \dots, \gamma_{R_{n-1}} \} < \gamma_{req}, \right. \\ \left. \& \gamma_{req} < \gamma_{R_n} < \gamma_0 \right] + \Pr \left(\gamma_{req} \leq \gamma_{R_1} < \gamma_0 \right), \gamma_0 \geq \gamma_{req}, \\ \Pr \left[\max \{ \gamma_{R_1}, \gamma_{R_2}, \dots, \gamma_{R_N} \} < \gamma_0 \right], \gamma_0 < \gamma_{req}. \end{cases} \quad (44)$$

Since all the links between S and R_n undergo i.i.d. SR fading, with the help of [51], (44) can be further expressed as

$$Z_1 = \begin{cases} [F_{\gamma_{R_n}}(\gamma_{req})]^N + \sum_{n=2}^N \left\{ [F_{\gamma_{R_n}}(\gamma_{req})]^{n-1} \right. \\ \left. \times [F_{\gamma_{R_n}}(\gamma_0) - F_{\gamma_{R_n}}(\gamma_{req})] \right\} \\ + [F_{\gamma_{R_1}}(\gamma_0) - F_{\gamma_{R_1}}(\gamma_{req})], \gamma_0 \geq \gamma_{req} \\ [F_{\gamma_{R_n}}(\gamma_0)]^N, \gamma_0 < \gamma_{req}. \end{cases} \quad (45)$$

After some mathematical derivation and simplification, (45) can be rewritten as

$$Z_1 = \begin{cases} 1 - \sum_{n=0}^{N-1} [F_{\gamma_{R_n}}(\gamma_{req})]^n [1 - F_{\gamma_{R_n}}(\gamma_0)], \gamma_0 \geq \gamma_{req}, \\ [F_{\gamma_{R_n}}(\gamma_0)]^N, \gamma_0 < \gamma_{req}. \end{cases} \quad (46)$$

To further derive the above expressions, it is essential to obtain the closed-form expression of $F_{\gamma_{R_n}}(\gamma_0)$. According to (8), (9), and (14), we will obtain

$$F_{\gamma_{R_n}}(\gamma_0) = \underbrace{F_{\gamma_{R_n x_1}}(\gamma_0)}_{Z_{11}} + \underbrace{F_{\gamma_{R_n x_2}}(\gamma_0)}_{Z_{12}} - \underbrace{F_{\gamma_{R_n x_1}}(\gamma_0)}_{Z_{11}} \underbrace{F_{\gamma_{R_n x_2}}(\gamma_0)}_{Z_{12}}. \quad (47)$$

By substituting (23) and (26) into (8) and (9), we will further obtain

$$F_{\gamma_{R_n x_1}}(\gamma_0) = \Pr(\lambda_{S R_n} \leq G_1 \lambda_{J_1 R_n} + G_2) \\ = \int_0^\infty F_{\lambda_{S R_n}}(G_1 \lambda_{J_1 R_n} + G_2) f_{\lambda_{J_1 R_n}}(y) dy, \quad (48)$$

where $F = \frac{\gamma_0}{a_1^2 - \gamma_0(a_2^2 + k_{s r_n}^2)}$, $G_1 = (1 + k_{j_1 r_n}^2) F$, $G_2 = E \times F$.

For notational convenience, a new function $\mathcal{X}(\alpha, \beta, \gamma)$ is defined, which is expressed as

$$\mathcal{X}(\alpha, \beta, \gamma) = 1 - \sum_{k_1=0}^{m_{S R_n}-1} \dots \sum_{k_M=0}^{m_{S R_n}-1} \sum_{p=0}^{\Lambda_{S R_n}-1} \frac{\Xi(M) (\Lambda_{S R_n} - 1)! \bar{\lambda}_{J_1 R_n}^{-J_1} \alpha^p}{p! \Delta_{S R_n}^{\Lambda_{S R_n}-p} (J_1 - 1)!} \\ \times \sum_{q=0}^p \binom{p}{q} \left(\frac{E}{1 + k_{j_1 r_n}^2} \right)^q \frac{\Gamma(p - q + J_1)}{\gamma^{p-q+J_1}} e^{-\Delta_{S R_n} \beta}. \quad (49)$$

After some trivial computations, $F_{\gamma_{R_n x_1}}(\gamma_0)$ and $F_{\gamma_{R_n x_2}}(\gamma_0)$ are finally given as $Z_{11} = \mathcal{X}(G_1, G_2, G_3)$ and $Z_{12} = \mathcal{X}(G_4, G_5, G_6)$, where $G_3 = \Delta_{S R_n} G_1 + 1/\lambda_{J_1 R_n}$, $F_1 = \frac{\gamma_0}{a_2^2 - \gamma_0 k_{s r_n}^2}$, $G_4 = F_1 (1 + k_{j_1 r_n}^2)$, $G_5 = F_1 \times E$, and $G_6 = \Delta_{S R_n} G_4 + 1/\lambda_{J_1 R_n}$. Then, by substituting Z_{11} and Z_{12} into (47), $F_{\gamma_{R_n}}(\gamma_0)$ will be obtained. Actually, it is not difficult to find that $F_{\gamma_{R_n}}(\gamma_0)$ is the CDF of γ_{R_n} . Thus, by taking (47) into (46) and following some mathematical steps, Z_1 will be obtained.

Next, we derive the analytical expression for Z_2 , which can be separated into several components as shown below

$$\begin{aligned} Z_2 &= \Pr(\min[\min(\gamma_{D_1x_1}, \gamma_{D_2x_1}), \gamma_{D_2x_2}] \leq \gamma_0) \\ &= 1 - \underbrace{[1 - F_{\gamma_{D_1x_1}}(\gamma_0)]}_{Z_{21}} \underbrace{[1 - F_{\gamma_{D_2x_1}}(\gamma_0)]}_{Z_{22}} \underbrace{[1 - F_{\gamma_{D_2x_2}}(\gamma_0)]}_{Z_{23}}. \end{aligned} \quad (50)$$

For notational convenience, a new function $\mathcal{D}(\alpha, \beta, \gamma, \rho, v)$ is defined, which is expressed as

$$\begin{aligned} \mathcal{D}(\alpha, \beta, \gamma, \rho, v) &= \sum_{u=0}^{M-1} \frac{\rho^{-J_2}}{u!(J_2-1)!} \left(\frac{\alpha}{\beta}\right)^u e^{-\frac{\gamma}{\beta}} \\ &\times \sum_{v=0}^u \binom{u}{v} \left(\frac{\gamma}{\alpha}\right)^v \frac{\Gamma(u-v+J_2)}{v^{u-v+J_2}}. \end{aligned} \quad (51)$$

With the help of (51), each part in (50) can be expressed as $Z_{21} = \mathcal{D}(U_1, \bar{\lambda}_{R_n D_1}, U_2, \bar{\lambda}_{J_2 D_1}, U_3)$, $Z_{22} = \mathcal{D}(V_1, \bar{\lambda}_{R_n D_2}, V_2, \bar{\lambda}_{J_2 D_2}, V_3)$, $Z_{23} = \mathcal{D}(W_1, \bar{\lambda}_{R_n D_2}, W_2, \bar{\lambda}_{J_2 D_2}, W_3)$, and $U_1 = V_1 = \frac{\gamma_0(1+k_j^2)}{b_1^2 - \gamma_0(b_2^2 + A - 1)}$, $W_1 = \frac{\gamma_0(1+k_j^2)}{b_2^2 - \gamma_0(A - 1)}$, $U_2 = \frac{\gamma_0 C}{b_1^2 - \gamma_0(b_2^2 + A - 1)}$, $V_2 = \frac{\gamma_0 D}{b_2^2 - \gamma_0(b_2^2 + A - 1)}$, $W_2 = \frac{\gamma_0 D}{b_2^2 - \gamma_0(A - 1)}$, $U_3 = \frac{U_1}{\bar{\lambda}_{R_n D_1}} + \frac{1}{\bar{\lambda}_{J_2 D_1}}$, $V_3 = \frac{V_1}{\bar{\lambda}_{R_n D_2}} + \frac{1}{\bar{\lambda}_{J_2 D_2}}$, $W_3 = \frac{W_1}{\bar{\lambda}_{R_n D_2}} + \frac{1}{\bar{\lambda}_{J_2 D_2}}$. By taking the above expressions into (50), Z_2 will be obtained. Then by taking (46) and (50) into (33), $P_{out}(\gamma_0)$ is obtained.

APPENDIX C PROOF OF THEOREM 3

Recalling (34), it can be divided into two components, denoted as R_1 and R_2 , which correspond to the CTR for decoding x_1 and x_2 , respectively. Accordingly, the overall CTR of the considered system is obtained by adding these two terms. Then, we first derive R_1 as

$$\begin{aligned} R_1 &= E[\log_2(1 + \gamma_1)] = \int_0^{\gamma_{\max_1}} \log_2(1 + x) f_{\gamma_1}(x) dx \\ &= \log_2(1 + x) F_{\gamma_1}(x) \Big|_0^{\gamma_{\max_1}} - \int_0^{\gamma_{\max_1}} \frac{F_{\gamma_1}(x)}{\ln 2(1 + x)} dx, \end{aligned} \quad (52)$$

where $\gamma_1 = \min(\gamma_{D_2x_1}, \gamma_{D_1x_1})$, $\gamma_{\max_1}$ is the maximal value of γ_1 which can be derived from (10) and (12) with value $\gamma_{\max_1} = \frac{b_1^2}{b_2^2 + A - 1}$. To proceed, the CDF of γ_1 is required, where $F_{\gamma_1}(x)$ can be expressed as

$$F_{\gamma_1}(x) = 1 - Z_{21}Z_{22}. \quad (53)$$

It should be noted that the integral term in (52) is difficult to evaluate using conventional analytical methods. Therefore, with the help of formula (25.4.30) in [50], the original interval is first transformed into the corresponding standard interval. Then, by substituting the zeros of the Legendre polynomials and the corresponding weight factors provided in Table (25.4) in [50], we have

$$R_1 = \log_2(1 + \gamma_{\max_1}) F_{\gamma_1}(\gamma_{\max_1}) - \sum_{i=1}^n \frac{\omega_i \gamma_{\max_1} F_{\gamma_1}(x_i)}{2 \ln 2(1 + x_i)}. \quad (54)$$

Next, by following a similar way, with the help of (11), we can obtain

$$\begin{aligned} R_2 &= E[\log_2(1 + \gamma_{D_2x_2})] \\ &= \log_2(1 + x) F_{\gamma_{D_2x_2}}(x) \Big|_0^{\gamma_{\max_2}} \\ &\quad - \int_0^{\gamma_{\max_2}} \frac{F_{\gamma_{D_2x_2}}(x)}{\ln 2(1 + x)} dx. \end{aligned} \quad (55)$$

Since the CDF of $\gamma_{D_2x_2}$ can be written as $1 - Z_{23}$, then the expression of R_2 is given by

$$\begin{aligned} R_2 &= \log_2(1 + \gamma_{\max_2}) F_{\gamma_{D_2x_2}}(\gamma_{\max_2}) - \sum_{i=1}^n \frac{\omega_i \gamma_{\max_2} F_{\gamma_{D_2x_2}}(x_i)}{2 \ln 2(1 + x_i)}. \end{aligned} \quad (56)$$

It is worth noting that the integrals in (52) and (55) are analytically intractable in closed form. Therefore, by leveraging the Gauss-Legendre quadrature method as in [50], and omitting intermediate steps, the final expression of (34) can be obtained by adding (54) and (56).

APPENDIX D PROOF OF THEOREM 4

Before conducting the asymptotic analysis, we first derive the high SNR asymptotic expressions for each channel. According to [39], the PDF of λ_{SR_n} at high SNR can be expressed as

$$f_{\lambda_{SR_n}}(x) = \frac{\alpha_{SR_n}^M x^{M-1}}{(M-1)! \bar{\lambda}_{SR_n}^M} + o(x^M). \quad (57)$$

From [39], when $\bar{\lambda}_o = \bar{\lambda}_e$, $o = 1, 2, \dots, O$, (26) can be rewritten as

$$f_{\lambda_e}(x) = \frac{\bar{\lambda}_e^{-O}}{(O-1)!} x^{O-1} e^{-\frac{x}{\bar{\lambda}_e}}. \quad (58)$$

Furthermore, the CDF of $\lambda_{R_n D_i}$ at high SNR is given by

$$F_{\lambda_{R_n D_i}}(x) = \frac{1}{M!} \left(\frac{x}{\bar{\lambda}_{R_n D_i}} \right)^M. \quad (59)$$

By using a similar method to obtain the $P_{out}(\gamma_0)$, $P_{out}(\gamma_0)$ is divided into two parts, namely Z_1^∞ and Z_2^∞ . Here, we first present the Z_1^∞ , the same as (44), which can be expressed by

$$Z_1^\infty = \begin{cases} 1 - \sum_{n=0}^{N-1} \left[F_{\gamma_{R_n}}^\infty(\gamma_{req}) \right]^n \left[1 - F_{\gamma_{R_n}}^\infty(\gamma_0) \right], & \gamma_0 \geq \gamma_{req}, \\ \left[F_{\gamma_{R_n}}^\infty(\gamma_0) \right]^N, & \gamma_0 < \gamma_{req}. \end{cases} \quad (60)$$

According to (47), $F_{\gamma_{R_n}}^\infty(\gamma_0)$ can be further expressed by $F_{\gamma_{R_n}}^\infty(\gamma_0) = Z_{11}^\infty + Z_{12}^\infty - Z_{11}^\infty Z_{12}^\infty$. For notational convenience, two new functions are defined, which are expressed as

$$\begin{aligned} S(\alpha, \beta, \gamma) &= \frac{\alpha^M}{M!} \sum_{z=0}^M \binom{M}{z} \left(\frac{\beta}{\alpha} \right)^z \frac{\gamma^{-J_2}}{(J_2-1)!} \frac{\Gamma(M-z+J_2)}{\gamma^{-M+z-J_2}}. \end{aligned} \quad (61)$$

$$\mathcal{Y}(\alpha, \beta) = \frac{(\alpha_{SR_n} \alpha)^M}{M!} \frac{\bar{\lambda}_{J_1 R_n}^{-J_1}}{(J_1 - 1)!} \sum_{z=0}^M \left(\frac{\beta}{\alpha} \right)^z \times \binom{M}{z} \frac{\Gamma(M - z + J_1)}{\bar{\lambda}_{J_1 R_n}^{-M+z-J_1}}. \quad (62)$$

When calculating Z_{11}^∞ , by taking (57) and (58) into (48), we will obtain $Z_{11}^\infty = \frac{\mathcal{Y}(G_1, G_2)}{\bar{\lambda}_{SR_n}^M}$. By a similar way, Z_{12}^∞ can be further expressed as $Z_{12}^\infty = \frac{\mathcal{Y}(G_4, G_5)}{\bar{\lambda}_{SR_n}^M}$. Then, by substituting Z_{11}^∞ and Z_{12}^∞ into $F_{\gamma_{R_n}}^\infty(\gamma_0)$, the expression for $F_{\gamma_{R_n}}^\infty(\gamma_0)$ is obtained. Here, $F_{\gamma_{R_n}}^\infty(\gamma_0)$ also represents the CDF of γ_{R_n} . Next, replacing γ_0 with γ_{req} in this expression yields $F_{\gamma_{R_n}}^\infty(\gamma_{req})$. Finally, by applying the proposed threshold-based link selection scheme, the value of Z_1^∞ can be determined.

Next, we focus on Z_2^∞ , which is consisted as Z_{21}^∞ , Z_{22}^∞ , and Z_{23}^∞ , expressing as

$$Z_2^\infty = 1 - Z_{21}^\infty Z_{22}^\infty Z_{23}^\infty. \quad (63)$$

In a similar way, Z_2^∞ is also composed of the three parts, which are shown as $Z_{21}^\infty = \mathcal{L}(U_1, \bar{\lambda}_{R_n D_1}, U_2, \bar{\lambda}_{J_2 D_1})$, $Z_{22}^\infty = \mathcal{L}(V_1, \bar{\lambda}_{R_n D_2}, V_2, \bar{\lambda}_{J_2 D_2})$, $Z_{23}^\infty = \mathcal{L}(W_1, \bar{\lambda}_{R_n D_2}, W_2, \bar{\lambda}_{J_2 D_2})$, and $\mathcal{L}(\alpha, \beta, \gamma, \rho) = 1 - \frac{1}{M!} \left(\frac{\alpha}{\beta} \right)^M \sum_{z=0}^M \binom{M}{z} \left(\frac{\gamma}{\alpha} \right)^z \frac{\rho^{-J_2}}{(J_2 - 1)!} \frac{\Gamma(M - z + J_2)}{\rho^{-M-z+J_2}}$. Then, by inserting the above expressions into (63), Z_2^∞ will be obtained.

To further analyze the diversity order and array gain of the proposed system more effectively, we assume $\bar{\lambda}_{R_n D_i} = \bar{\lambda}_{SR_n} = \bar{\lambda}$. With this assumption and neglecting higher-order terms, all the above expressions can be rewritten as

$$P_{out}^\infty(\gamma_0) = \begin{cases} [(Q_1 + Q_2) + (P_1 + P_2 + P_3)] \left(\frac{1}{\bar{\lambda}} \right)^M, & \gamma_0 \geq \gamma_{req}, \\ (P_1 + P_2 + P_3) \left(\frac{1}{\bar{\lambda}} \right)^M, & \gamma_0 < \gamma_{req}, \end{cases} \quad (64)$$

where $P_1 = \mathcal{S}(U_1, U_2, \bar{\lambda}_{J_2 D_1})$, $P_2 = \mathcal{S}(V_1, V_2, \bar{\lambda}_{J_2 D_2})$, $P_3 = \mathcal{S}(W_1, W_2, \bar{\lambda}_{J_2 D_2})$, $Q_1 = \mathcal{Y}(G_1, G_2)$, $Q_2 = \mathcal{Y}(G_4, G_5)$. Then, by taking the above expressions into (64), we will further obtain the diversity order and array gain, which are shown in (36). The proof is completed.

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