



Active IRS-Enhanced UAV Relaying for Secure Hybrid RF/FSO Communications: Joint Phase Shift, Beamforming, and Power Optimization

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ABSTRACT

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We investigate an unmanned aerial vehicle (UAV)-assisted wireless communication network that employs intelligent reflecting surfaces (IRS) to enhance physical-layer security and reliability against eavesdropping. The proposed system integrates a cooperative decode-and-forward (DF) relay realized by a UAV with hybrid radio frequency/free-space optical (RF/FSO) links to improve outage performance while securing data transmission. An IRS (passive or active) is deployed on the UAV to reconfigure wireless propagation, creating virtual line-of-sight paths and directing signal energy toward legitimate users. We provide a simulation-based numerical analysis of the outage probability and secrecy performance of the proposed system under practical impairments including imperfect channel state information (CSI), IRS phase errors, and FSO pointing misalignment. Artificial noise (AN) and joint power/beamforming optimization are employed to enhance secrecy performance. Simulation-based results in the downlink show that aerial IRS with many elements significantly reduces secrecy outage, and active IRS can not only compensate for multiplicative path loss but also achieve substantial secrecy-capacity gains.

1. INTRODUCTION

Driven by the development of sixth-generation (6G) networks, the need for secure, energy-efficient, and high-throughput wireless communications has become more urgent. Two attractive candidates to realize such a vision are the unmanned aerial vehicles (UAVs) and intelligent reflecting surfaces (IRSs), which have become more widely acknowledged as enablers of this paradigm. UAVs provide flexible relaying with adaptive positioning on demand, whereas IRSs alter the propagation environment by way of passive reflection. Their combination has the potential to bring not only a significant performance enhancement on coverage and capacity but also to facilitate the enhancement of physical layer security (PLS), and hence UAV-IRS systems are seen as prospective pillars of future wireless architectures. In particular, the high-quality line-of-sight channels UAVs can provide are favorable not only to legitimate users but also to eavesdroppers, which can be resolved by reconfigurable intelligent surfaces through smartly modifying the wireless environment to boost secure transmission [1, 2].

However, practical implementation of IRS-aided systems is not without challenges, such as the “double fading” effect, which can be mitigated by active IRS and relaying protocols [3]. Active IRS introduces controllable power amplifiers into its reflecting elements to compensate for the path loss,

enabling dynamic user scheduling according to channel quality, and the best achievable secrecy capacity can be obtained [4].

The latest studies reveal that UAVs carrying IRSs can enhance beneficial links and reduce information leakage to the eavesdroppers. For example, IRS-enabled UAVs in low-altitude platform and vehicular IoT paradigms mitigate secrecy outage probability in high mobility scenarios [5]. Cognitive UAV communications with IRS assistance are further investigated to achieve secure spectrum sharing, where an iterative optimization solution enhances secrecy capacity without violating the interference limitations [6]. Safe transmission protocols for IRS-UAV-enabled was investigated as well [7], and the design of a multi-functional RIS mounted on a UAV to defeat jamming and multipath fading has been studied, which enhances the security of wireless aerial networks [8].

The synergistic attributes of the radio frequency (RF) and free-space optical (FSO) systems have also been intensively investigated. Although FSO provides high-capacity line-of-sight communication, it is sensitive to atmospheric turbulence and pointing errors. RF links, even if much smaller in bandwidth, are known to be more resilient to blockage and turbulence. Hybrid IRS-assisted radio frequency/free-space optical (RF/FSO) systems are therefore a good candidate to exploit these trade-offs. Recent work illustrates that the

secrecy outage probability and the achievable secrecy capacity in such systems significantly depend on turbulence parameters, complexity of IRS phase alignment, as well as UAVs' position, and, hence, these results demonstrate the

importance of performing the joint performance analysis [9-11]. A structured comparison between the proposed system and the most relevant related works is summarized in Table 1.

Table 1. Comparison with related work

Reference	IRS Type	UAV Role	Channel	Relay	CSI	Method	Optimized Variables	AN	Metrics	Scope Limitation
[1]	Passive	UAV receiver	RF	Direct	Perfect	Alternating optimization	Uplink power, UAV trajectory, RIS phase	No	Average secrecy rate	RF uplink; no DF or FSO
[2]	Active	UAV carries active RIS; separate DF relays	RF	DF	Statistical	Analytical OP/IP	Relay selection, RIS placement	No	Outage, intercept probability, SRT	No FSO, AN, or joint beamforming
[4]	Active	UAV transmitter; fixed active IRS	RF	Direct	Perfect	BCD + SCA/CCT/MM	User scheduling, UAV trajectory, beamforming, active-IRS coefficients	No	Average secrecy capacity	RF-only; fixed IRS; no DF or FSO
[5]	Passive	Low-altitude platform	RF	Direct	Perfect	Convex optimization	Power, phase shifts	No	Secrecy capacity	DF hybrid RF/FSO relaying not considered
[9]	Passive	No UAV	RF	DF relay + IRS	Perfect	Power-allocation analysis	Power allocation	No	Security and reliability	No FSO, active IRS, UAV, or imperfect CSI
[10]	Passive	No UAV	FSO→RF	Dual-hop relay	Perfect	Analytical SOP	NOMA power and RIS scenarios	No	Secrecy outage probability	No active IRS, UAV, AN, or joint beamforming
[12]	Hybrid active/passive	UAV-mounted hybrid IRS	RF	Direct	Perfect	SCA/FP/MM	Beamforming; active/passive phase and amplitude	No	Achievable rate	No DF, FSO, AN, or secrecy-outage analysis
This work	Active/passive	UAV DF relay + IRS carrier	RF→FSO	DF	Imperfect	Phase alignment + space AN + grid search	IRS phase/gain, beamforming, power split	Yes	Secrecy rate, outage, sensitivity	Fixed geometry; normalized simulation model

Note: IRS = intelligent reflecting surfaces; UAV = unmanned aerial vehicle; CSI = channel state information; AN = artificial noise; RF = radio frequency; DF = decode-and-forward; FSO = free-space optical; BCD = block coordinate descent; SCA = successive convex approximation.

Energy-efficient design and energy harvesting for UAV-IRS systems have been reported in studies [12, 13]. Although AI/RL-based techniques (e.g., DRL for beamforming and phase-shift control in UAV-IRS networks [12]) have also been reported, they are outside the scope of the present work, which relies on optimization-based BCD-SCA rather than learning-based methods.

While most works are dedicated to passive IRSs, there exist a few results on active IRS architectures, where each reflecting element is equipped with a low-power amplifier. The multiplicative fading loss inherent in cascaded channels is compensated for by active IRSs, which also significantly enhance the secrecy rate and decrease the outage probability. UAV-borne active IRS relays have also been demonstrated to have better performance over their passive counterparts, especially under the circumstance of strong turbulence or weak line-of-sight links [4].

However, these are not yet been overcome. Various works consider perfect channel state information (CSI) and ideal IRS phase shift under the mobility of UAV and dynamic FSO turbulence. Combined characterization of outage probability and secrecy rate under UAV-IRS hybrid RF/FSO networks

remains scarce. Analysis on the effect of practical impairments, including IRS phase quantization, UAV trajectory uncertainty, atmospheric turbulence, and hardware imperfections, on secure outage performance is deepened. Also, the integration of cooperative decode-and-forward (DF) relaying and artificial noise (AN) injection in a UAV-mounted IRS system has not been fully studied under imperfect CSI conditions.

Inspired by these challenges, we investigate in this paper a complete outage probability and secrecy analysis of UAV-aided hybrid RF/FSO systems with the assistance of the IRS technique. We introduce a cooperative DF scheme where the UAV-mounted IRS simultaneously enhances the legitimate link and weakens the eavesdropper's channel. The system explicitly considers turbulence-induced fading, pointing errors, Nakagami-m RF fading, and imperfect CSI. Extensive simulations illustrate that UAV-assisted IRS outperforms UAV-/IRS-assisted secure communication significantly, confirming its feasibility in 6G secure communication networks.

More directly related to this work, recent studies have focused on: (i) active IRS architectures with element-level

amplification for compensating multiplicative path loss and enhancing secrecy [1, 2]; (ii) FSO pointing errors and Gamma-Gamma turbulence modeling in secure hybrid RF/FSO systems [9]; (iii) DF-based secure relaying with AN injection [5]; and (iv) secure IRS designs under imperfect CSI, where estimation errors reduce both beamforming quality and AN orthogonality [10]. The proposed manuscript integrates these directly relevant aspects into a single joint-optimization framework. The main contributions of this paper are as follows:

1. **Novel System Architecture:** We introduce active IRS, cooperative DF relaying, and hybrid RF/FSO links with artificial noise injection to the UAV-assisted secure communication system model for better physical-layer security.

2. **Accurate Channel Modeling:** A tight channel model is developed, which accounts for Nakagami-m RF fading, Gamma-Gamma ATM for FSO, pointing errors, practical impairments such as imperfect CSI, IRS phase quantization errors, and noise from amplifiers embedded in active IRS elements.

3. **Joint Optimization Framework:** A non-convex joint optimization problem based on IRS phase shifts (Θ), beamforming vectors (G), and power allocation factor (α) to maximize the average secrecy capacity is formulated and solved. A computationally efficient block coordinate descent (BCD) scheme combined with SCA and ternary search is proposed, while convergence is also analyzed.

4. **Active vs. Passive IRS Analysis:** reports a comparative study on passive and active IRS-based UAV-mounted hybrid RF/FSO secure communications, revealing that active IRS is capable of compensating for the multiplicative path loss and achieves up to 185% secrecy-capacity enhancement over baseline schemes under an equal-transmit-power comparison, and about 97% under a stricter equal-total-power comparison in which the amplifier supply power is fully accounted for; both scenarios are reported based on 10^4 Monte Carlo realizations, guidelines for the deployment, the optimal UAV altitude (60 m), the optimal IRS array size ($N = 128$ elements), and the optimal power allocation ($\alpha^* \approx 0.59$) are obtained, trading off secrecy capacity against computational complexity.

5. **Performance characterization:** We provide an analytical and numerical analysis of the secrecy outage probability under practical conditions, showing that the proposed system has $P_{out} = 10^{-3}$ at 12 dB SNR (which corresponds to 10 dB gain as compared with the baseline system) with 4.82 bits/s/Hz of secrecy capacity.

The remainder of this paper is organized as follows. Section 2 presents the system model, problem formulation, and joint optimization algorithms. Section 3 provides simulation results and performance analysis. Section 4 concludes the paper with future research directions.

2. METHODOLOGY

2.1 System and channel model

We consider a UAV-assisted secure communication network illustrated in Figure 1, where a ground source (S) transmits confidential information to a legitimate destination (Bob, B) via a UAV-mounted IRS relay (R), while an eavesdropper (Eve, E) attempts to intercept the transmission.

The system employs a hybrid RF/FSO architecture operating in two distinct phases. In Phase 1, the S→R link operates over RF (Nakagami-m fading) to ensure reliable signal reception at the UAV relay under mobility conditions. In Phase 2, the R→B link operates over FSO (Gamma-Gamma atmospheric turbulence with pointing errors) to achieve high-capacity secure transmission to Bob. DF decoding takes place entirely at the UAV relay at the end of Phase 1: the UAV fully decodes the received signal before re-encoding and retransmitting in Phase 2. In the active IRS mode, each of the N reflecting elements phase-shifts and amplifies the impinging signal using an embedded low-noise amplifier, with element amplitude gain α_n , where in the passive IRS mode, each element only phase-shifts the signal, with unit amplitude gain α_n . Eve is assumed capable of intercepting transmissions in both phases, representing the worst-case eavesdropping scenario. In details, the UAV hovers at altitude h_R and is equipped with an IRS comprising N reconfigurable reflecting elements. The source-to-relay link experiences Nakagami- m with channel coefficient; $h_{SR} = \sqrt{\beta_{SR}} \cdot g_{SR}$, $g_{SR} \sim \text{Nakagami}(m_{SR}, \Omega_{SR})$ where $\beta_{SR} = \rho_0(d_{SR}/d_0)^{-\alpha_{RF}}$ represents large-scale path loss with reference distance $d_0 = 1$ m, path loss exponent $\alpha_{RF} = 2.5$, and ρ_0 is the reference path gain. Then the instantaneous channel gain follows $|h_{SR}|^2 \sim \text{Gamma}(m_{SR}, \frac{\Omega_{SR}}{m_{SR}})$. The relay-to-destination FSO link is characterized by atmospheric turbulence modeled via Gamma-Gamma distribution and pointing errors. The channel coefficient is:

$$h_{RB}^{\text{FSO}} = h_a \cdot h_p \cdot h_l \quad (1)$$

where, $h_a \sim \text{Gamma-Gamma}(\alpha, \beta)$ is the atmospheric turbulence following a Gamma-Gamma distribution with shape parameters α and β (moderate turbulence, Rytov variance $\sigma^2_R = 0.5$), with PDF [14].

$$f_{h_a}(x) = \frac{2(\alpha\beta)^{(\alpha+\beta)/2}}{\Gamma(\alpha)\Gamma(\beta)} x^{(\alpha+\beta)/2-1} K_{\alpha-\beta}(2\sqrt{\alpha\beta}x), x > 0 \quad (2)$$

The pointing error; $h_p = A_0 \exp\left(-\frac{2r^2}{w_{eq}^2}\right)$, where r is radial displacement, w_{eq} is equivalent beam radius, and $A_0 = [\text{erf}(v)]^2$ with $v = \sqrt{\pi}a/(\sqrt{2}w_{eq})$. The pointing error coefficient follows $|h_p|^2 \sim \text{Rayleigh}(\xi^2)$, $\xi^2 = \frac{w_{eq}^2}{4\sigma_s^2}$ where σ_s is the jitter standard deviation at the receiver. Besides, the path loss, $h_l = \exp(-\sigma_{FSO}d_{RB})$ accounts for absorption/scattering over distance d_{RB} (relay-to-Bob distance) with attenuation coefficient σ_{FSO} (dB/km; $\sigma_{FSO} = 0.43$ dB/km for clear sky, 4.2 dB/km for light fog, 20 dB/km for moderate fog). Eavesdropper channels are modeled similarly as $h_{SE} \sim \text{Nakagami}(m_{SE}, \Omega_{SE})$, $h_{RE}^{\text{FSO}} \sim \text{Gamma-Gamma}(\alpha_E, \beta_E)$ with typically $m_{SE} < m_{SR}$ and $\alpha_E < \alpha$, reflecting Eve's disadvantaged position.

The IRS phase shift matrix is defined as:

$$\Theta = \text{diag}(\theta_1, \theta_2, \dots, \theta_N), \theta_n = \alpha_n e^{j\phi_n} \quad (3)$$

where, α_n as the element gain: passive IRS (unit gain, phase-shift only); $\alpha_n = 1$; and active IRS: $\alpha_n = \sqrt{G_n}$ (amplification gain, where G_n is the element power gain; e.g., $G_n = 4$ gives $\alpha_n = 2$, equivalent to 6 dB), $\phi_n \in [0, 2\pi]$.

Then, the cascaded channel from source to Bob through IRS

is formulated as:

$$h_{eff}^B = h_{RB}^T \Theta G h_{SR} \quad (4)$$

where, $h_{SR} \in \mathbb{C}^{N \times 1}$, is channel vector from S to IRS elements. $G = \text{diag}(g_1, \dots, g_N)$ is the beamforming weight matrix at UAV, and $h_{RB} \in \mathbb{C}^{N \times 1}$ is the channel vector from IRS elements to Bob.

Similarly, the effective eavesdropper channel is:

$$h_{eff}^E = h_{RE}^T \Theta G h_{SE} \quad (5)$$

To this end, the received signals at the UAV relay in the first phase (S→R) can be formulated as follows:

$$y_R = \sqrt{P_S} h_{SR} x + n_R \quad (6)$$

where, x is the unit-power information symbol, P_S is source transmit power ($P_S = P_T = 1$ W normalized), and $n_R \sim \mathcal{CN}(0, \sigma_R^2)$ is the additive wight gaussian noise. Here σ_R^2 , σ_B^2 , and σ_E^2 denote the noise powers at relay, Bob, and Eve respectively. In the second phase (R→B with artificial noise [15]), the relay employs DF with power splitting [16]. The transmitted signal is:

$$s_R = \sqrt{\alpha P_R} G \hat{x} + \sqrt{(1 - \alpha) P_R} w_{AN} z \quad (7)$$

where, $\alpha \in (0,1)$ is the power allocation factor, P_R is the relay transmit power, $\hat{x}$ is the decoded symbol, w_{AN} is the artificial noise beamforming vector, $|w_{AN}|^2 = 1$, and $z \sim \mathcal{CN}(0,1)$ is the artificial noise.

To this end, Bob's received signal can be obtained as:

$$y_B = h_{eff}^B s_R + n_B = \sqrt{\alpha P_R} h_{eff}^B G \hat{x} + \sqrt{(1 - \alpha) P_R} h_{eff}^B w_{AN} z + n_B \quad (8)$$

Thus, the Bob SINR can be computed as:

$$\gamma_B = \frac{\alpha P_R |h_{eff}^B G|^2}{(1 - \alpha) P_R |h_{eff}^B w_{AN}|^2 + \sigma_B^2} \quad (9)$$

assuming w_{AN} is designed to lie in the null-space of the estimated effective channel to Bob ($h_{eff}^B \perp w_{AN}$ under perfect CSI, so ideally $h_{eff}^B w_{AN} \approx 0$). The signal received at Eve can be calculated as follows:

$$y_E = h_{eff}^E s_R + n_E = \sqrt{\alpha P_R} h_{eff}^E G \hat{x} + \sqrt{(1 - \alpha) P_R} h_{eff}^E w_{AN} z + n_E \quad (10)$$

Similarly, the Eve SINR can be calculated as:

$$\gamma_E = \frac{\alpha P_R |h_{eff}^E G|^2}{(1 - \alpha) P_R |h_{eff}^E w_{AN}|^2 + \sigma_E^2} \quad (11)$$

Finally, the instantaneous secrecy rate under DF is obtained by:

$$R_s = [\min\{R_{SR}, R_{RB}\} - R_E]^+ \quad (12)$$

where, $R_{SR} = \log_2(1 + \gamma_{SR})$, is S→R capacity with $\gamma_{SR} = P_S |h_{SR}|^2 / \sigma_R^2$, the R→B capacity calculated by using $R_{RB} = \log_2(1 + \gamma_B)$, and the eavesdropper rate over both phases (worst-case scenario where Eve intercepts both the S→E link in Phase 1 and the R→E link in Phase 2) is defined by the worst hop as: $R_E = \max\{R_E^1, R_E^2\}$, where $R_E^1 = \log_2(1 + \gamma_E^1)$ is Eve's rate in Phase 1 and $R_E^2 = \log_2(1 + \gamma_E^2)$ is Eve's rate in Phase 2; finally, $[x]^+ = \max(0, x)$. Following DF relaying principles, the end-to-end achievable rate between S and Bob is limited by the weaker hop: $R_{SB} = \min\{R_{SR}, R_{RB}\}$. Therefore, the instantaneous secrecy rate R_s in Eq. (12) explicitly incorporates the DF bottleneck and worst-case eavesdropping over both phases. On the other hand, the average secrecy capacity can be computed by using:

$$C_s = E[R_s] \quad (13)$$

And the secrecy outage probability for target rate obtained by [17]:

$$P_{out} = \Pr(R_s < R_{th}) \quad (14)$$

2.2 Problem formulation

The goal is to maximize the average secrecy capacity of the two-hop UAV-IRS system while satisfying practical constraints. We formulate this as a joint optimization problem:

$$\max_{\theta, w, \alpha} C_s(\theta, w, \alpha) = E[R_s] \quad (15)$$

subject to:

$$\alpha P_T + (1 - \alpha) P_T \leq P_T \quad (15a)$$

$$|\varphi_n| \leq \varphi_m \alpha x, \forall n = 1, 2, \dots, N \quad (15b)$$

$$0 < \alpha < 1 \quad (15c)$$

where, $w \in \mathbb{C}^N$ is the beamforming weight vector at the UAV. The constraint Eq. (15a) ensures total power conservation at the relay (for passive IRS: $P_S + P_R = P_T$; for active IRS: $P_S + P_R + N \cdot P_{amp} \leq P_{budget}$, where $N \cdot P_{amp}$ is the amplifier supply power consumption, thereby enforcing a fair power comparison with the passive baseline. The optimization variables are coupled through the secrecy rate expression:

$$R_s = [\log_2(1 + \gamma_B) - \log_2(1 + \gamma_E)]^+ \quad (16)$$

where, γ_B and γ_E denote the SINR at the legitimate receiver Bob and Eavesdropper Eve, respectively.

2.3 Joint optimization via block coordinate descent

The formulated problem in Eqs. (15a)-(15c) is non-convex due to the coupled optimization variables in the cascaded channel model and the nonlinear SINR expressions. The coupling among different variables makes the direct solution of the original optimization problem challenging, as the variables cannot be optimized independently. We address this challenge using BCD combined with successive convex approximation (SCA), which has proven effective for similar IRS optimization problems [18, 19].

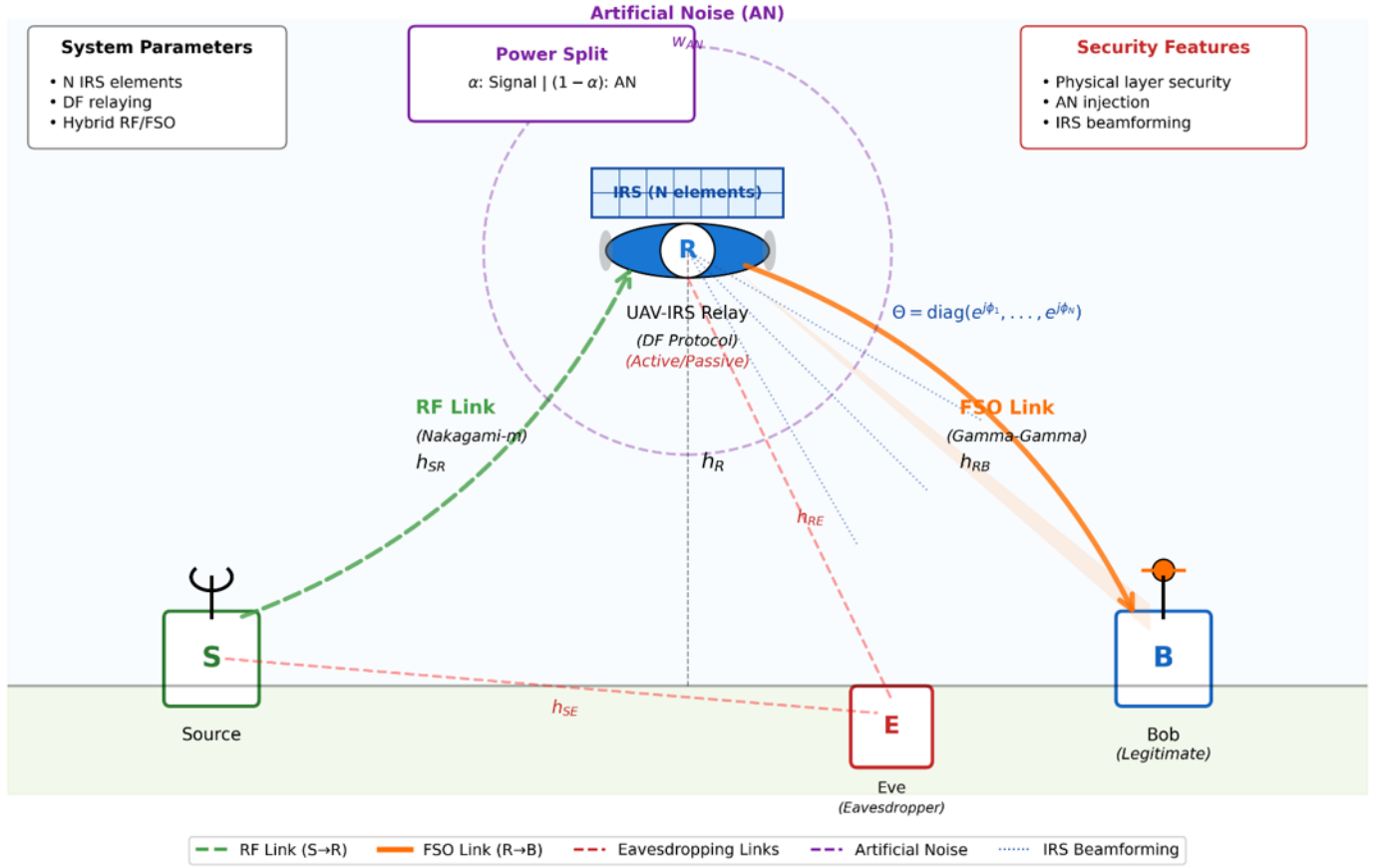


Figure 1. UAV-IRS assisted secure hybrid radio frequency/free-space optical (RF/FSO) communication system model

2.3.1 Block coordinate descent algorithm and convergence

BCD decomposes the joint optimization into three manageable subproblems, optimizing one variable while fixing the others. At each iteration k , we perform:

Step1:

$$\theta^{(k+1)} = \arg \max_{\theta} C_s(\theta, w^{(k)}, \alpha^{(k)}) \quad (17)$$

Step2:

$$w^{(k+1)} = \arg \max_w C_s(\theta^{(k+1)}, w, \alpha^{(k)}) \quad (18)$$

Step3:

$$\alpha^{(k+1)} = \arg \max_{\alpha} C_s(\theta^{(k+1)}, w^{(k+1)}, \alpha) \quad (19)$$

This procedure repeats until convergence, defined by:

$$\max(|\theta^{(k+1)} - \theta^{(k)}|, |w^{(k+1)} - w^{(k)}|, |\alpha^{(k+1)} - \alpha^{(k)}|) < \epsilon \quad (20)$$

where, $\epsilon = 10^{-4}$ is the convergence tolerance. Under the standard BCD-SCA assumptions: (i) each per-block subproblem admits a well-defined solution up to a controlled tolerance, and (ii) the SCA surrogate is a tight lower bound of the objective at the current iterate with matching gradients, the objective sequence generated by the proposed algorithm is monotonically non-decreasing and bounded and therefore converges. Because the subproblems are solved iteratively rather than to global optimality, and because the joint problem is non-convex, this convergence is to a stationary point of the

surrogate problem (in the KKT sense) rather than a global optimum, consistent with classical BCD-SCA analyses [20, 21]. Convergence to a stationary point of the original Lagrangian is therefore not claimed in the strict sense but is expected in practice; this is supported by the empirical monotonic convergence observed within a small number of iterations in our simulations.

2.3.2 Phase shift optimization via successive convex approximation

The IRS phase optimization subproblem is:

$$\max_{\theta} C_s(\theta, w, \alpha) \text{ s.t. } |\theta_n| \in [0, 2\pi], \forall n \quad (21)$$

This is non-convex due to the complex exponential coupling in the effective channel:

$$h_{\text{eff}} = h_{SI} + h_{ID} \odot (w \odot \theta) \quad (22)$$

where, $\odot$ denotes element-wise multiplication. We apply SCA by constructing a convex approximation of the secrecy capacity around the current solution $\theta^{(k)}$. The effective channel gain can be expressed as:

$$|h_{\text{eff}}|^2 = |h_{SI} + h_{ID} \odot (w \odot \theta)|^2 \quad (23)$$

First-order Taylor expansion around $\theta^{(k)}$ yields [22]:

$$|h_{\text{eff}}|^2 \approx |h_{\text{eff}}^{(k)}|^2 + 2\text{Re}\{(h_{\text{eff}}^{(k)})^* \nabla_{\theta} |h_{\text{eff}}|^2|_{\theta^{(k)}}\} \quad (24)$$

where the gradient is:

$$\nabla_{\theta} |h_{\text{eff}}|^2 = \text{conj}(h_{\text{ID}} \odot w) \odot (h_{\text{eff}}^{(k)})^* \quad (25)$$

The secrecy capacity is then approximated as:

$$C_s \approx \log_2 \left(1 + \frac{\alpha P_T \text{SNR} \cdot |h_{\text{eff}}^{(k)}|^2}{1 + \alpha P_T \text{SNR} \cdot |h_{\text{eff}}^{(k)}|^2 + \alpha P_T \text{SNR} \cdot |h_{\text{eve}}|^2} \right) - \log_2(1) \quad (26)$$

Projected gradient ascent update [23]:

$$\theta^{(k+1)} = \mathcal{P}(\theta^{(k)} + \mu_{\theta} \cdot \nabla_{\theta} C_s) \quad (27)$$

where, $\mu_{\theta} = 0.1$ is the step size and $\mathcal{P}(\cdot)$ projects the phases back to the unit circle:

$$\mathcal{P}(z) = \frac{z}{|z|} \quad (28)$$

To ensure monotonic improvement, we verify that $C_s(\theta^{(k+1)}) \geq C_s(\theta^{(k)})$. If not, we reduce the step size by half and retry. This SCA procedure repeats for $I_{\text{SCA}} = 5$ iterations per BCD iteration.

2.3.3 Beamforming vector optimization

The beamforming subproblem is:

$$\max_w C_s(\theta, w, \alpha) \text{ s.t. } |w| = 1 \quad (29)$$

The effective channel at Bob is:

$$h_{\text{eff,Bob}} = h_{\text{SI}} + h_{\text{ID}} \odot (\theta \odot w) \quad (30)$$

and the SINR becomes:

$$h_{\text{eff,Bob}} = h_{\text{SI}} + h_{\text{ID}} \odot (\theta \odot w) \quad (31)$$

$$\gamma_B = \alpha P_T \text{SNR} \cdot |h_{\text{eff,Bob}}|^2 = \alpha P_T \text{SNR} \cdot |w^H (h_{\text{ID}} \odot \theta \odot h_{\text{SI}})|^2 \quad (32)$$

Gradient Computation:

$$\nabla_w C_s = \frac{\partial C_s}{\partial \gamma_B} \cdot \nabla_w \gamma_B \quad (33)$$

where,

$$\nabla_w \gamma_B = 2 \cdot \alpha P_T \text{SNR} \cdot (h_{\text{ID}} \odot \theta) * (h_{\text{SI}})^T \quad (34)$$

Gradient Ascent with Normalization:

$$w^{(k+1)} = \frac{w^{(k)} + \mu_w \cdot \nabla_w C_s}{|w^{(k)} + \mu_w \cdot \nabla_w C_s|} \quad (35)$$

where, $\mu_w = 0.05$ is the beamforming step size. The normalization constraint $|w| = 1$ is maintained at each update, which can be interpreted as normalizing the antenna gains. This optimization runs for $I_w = 8$ iterations per BCD step.

2.3.4 Power allocation optimization via ternary search

The power allocation problem is:

$$\max_{0 < \alpha < 1} C_s(\theta, w, \alpha) \quad (36)$$

Key observation: The secrecy capacity $C_s(\alpha)$ is a unimodal function in α it increases to a unique maximum and then decreases. This property arises from the competing effects of signal power and artificial noise power on the secrecy rate: for small α , Bob's SINR is too low; for large α , the AN power is insufficient to sufficiently degrade Eve's SINR. While strict analytical unimodality of $C_s(\alpha)$ is difficult to establish in closed form under active-IRS noise, DF relaying, and imperfect CSI, we have verified it numerically over an extensive range of conditions. Specifically, we swept $\alpha \in [0, 1]$ in steps of 0.01 and computed $C_s(\alpha)$ at each of the operating points reported in the paper: SNR $\in \{-10, 0, 10, 20, 30\}$ dB, CSI quality $\sigma^2_{\text{CSI}} \in \{0, 0.01, 0.05, 0.1\}$, IRS size $N \in \{64, 128, 256\}$, and both passive and active IRS modes. In every tested scenario, $C_s(\alpha)$ exhibited a single global maximum with no observed local maxima. Under adverse conditions (e.g., $\sigma^2_{\text{CSI}} = 0.1$ combined with active IRS noise), the curve becomes flatter around the peak but remains unimodal. The unimodality assumption is therefore justified as an empirical property over the operating region of interest, and to exploit it, we employ ternary search, which converges to the global maximum in the unimodal regime in $O(\log(1/\epsilon))$ function evaluations.

This method guarantees convergence to the global optimum with precision 10^{-6} in approximately $\log_3(10^6) \approx 13$ iterations per BCD step [24].

2.4 Handling practical impairments

In realistic deployments, perfect CSI, phase accuracy, and hardware performance cannot be guaranteed. We explicitly model these impairments.

2.4.1 Channel state information

The estimated channel is modeled as [25]:

$$\hat{h} = h + \Delta h_{\text{CSI}} \quad (37)$$

where, $\Delta h_{\text{CSI}} \sim \mathcal{CN}(0, \sigma_{\text{CSI}}^2 |h|^2)$ represents Gaussian estimation error such that $\hat{h}_{\text{eff}}^B = h_{\text{eff}}^B + \Delta h_{\text{CSI}}$. Because the AN vector w_{AN} is designed in the null-space of the estimated channel $\hat{h}_{\text{eff}}^B$ rather than the true channel h_{eff}^B , a residual AN leakage appears at Bob under imperfect CSI: $|h_{\text{eff}}^B w_{\text{AN}}|^2 = |\Delta h_{\text{CSI}}^H w_{\text{AN}}|^2$, whose variance scales as $\sigma_{\text{leak}}^2 = \sigma_{\text{CSI}}^2 \cdot |h_{\text{eff}}^B|^2 \cdot |w_{\text{AN}}|^2$. This leakage term is explicitly added to Bob's SINR denominator, so that Bob's effective SINR under imperfect CSI is reduced from its perfect-CSI value ($\gamma_B^{\text{perfect}}$) to $\gamma_B^{\text{imp}} = \gamma_B^{\text{perfect}} / (1 + \eta_{\text{leak}} \cdot \sigma_{\text{CSI}}^2)$, with η_{leak} being the leakage sensitivity parameter. Consequently, a perfect null-space AN design is not assumed under imperfect CSI, and the residual leakage effect is quantified in all simulations reporting the imperfect-CSI results. In our simulations, we use three error levels:

Good CSI: $\sigma_{\text{CSI}}^2 = 0.01$ (high-quality feedback)

Moderate CSI: $\sigma_{\text{CSI}}^2 = 0.05$ (typical scenario)

Poor CSI: $\sigma_{\text{CSI}}^2 = 0.1$ (low feedback capacity)

The optimization in Algorithm 1 uses $\hat{h}$ instead of perfect h , and we analyze performance degradation across all three scenarios.

Algorithm 1. Ternary Search for Optimal α

1. Initialize: $\alpha_{\text{left}} = 0.1, \alpha_{\text{right}} = 0.99$
2. While $(\alpha_{\text{right}} - \alpha_{\text{left}}) > 10^{-6}$:
 - . Compute $m_1 = \alpha_{\text{left}} + (\alpha_{\text{right}} - \alpha_{\text{left}})/3$
 - . Compute $m_2 = \alpha_{\text{right}} - (\alpha_{\text{right}} - \alpha_{\text{left}})/3$
 - . Evaluate $C_s(m_1)$ and $C_s(m_2)$
 - . If $C_s(m_1) < C_s(m_2)$: set $\alpha_{\text{left}} = m_1$
 - . Else: set $\alpha_{\text{right}} = m_2$
3. Return: $\alpha^* = (\alpha_{\text{left}} + \alpha_{\text{right}})/2$

2.4.2 Intelligent reflecting surfaces phase quantization

Practical IRS elements cannot achieve arbitrary phase values due to hardware limitations. We model quantized phases as [26, 27].

$$\theta_{\text{quantized}} = \text{round}\left(\frac{\theta_{\text{optimal}}}{\Delta\phi}\right) \times \Delta\phi \quad (38)$$

where, $\Delta\phi = 2\pi/M$ and M is the number of quantization levels. We test $M \in 4, 8, 16$ bits per element and report performance vs. quantization resolution.

2.4.3 Active intelligent reflecting surfaces amplifier noise

For active IRS (amplification gain $\alpha_n = \sqrt{G_{\text{amp}}} = \alpha_n = 2$), each reflecting element includes a low-noise amplifier with gain $G_{\text{amp}} = 2$ and noise figure $NF = 4$ dB. The output of element n is [28, 29].

$$y_n = G_{\text{amp}}(h_{\text{in},n}x + n_{\text{amp},n}) \quad (39)$$

where, $n_{\text{amp},n} \sim \mathcal{CN}(0, \sigma_{\text{amp}}^2)$ with σ_{amp}^2 determined by the noise figure as $\sigma_{\text{amp}}^2 = (F - 1) \cdot k_B \cdot T \cdot B \cdot G_{\text{amp}}$, where $F = 10^{(NF/10)}$ is the noise figure in linear scale, k_B is Boltzmann's constant, $T = 290$ K is the operating temperature, and B is the signal bandwidth. This amplifier noise degrades Eve's SINR less than Bob's SINR (due to beamforming steering), but the effect is explicitly included in results. To properly account for the power cost of active IRS elements, we now define the total active-IRS power consumption as $P_{\text{IRS}} = N \cdot P_{\text{amp}}$, where P_{amp} is the per-element amplifier supply power ($P_{\text{amp}} \leq P_{\text{max}} = 10$ mW per element in our study). The overall system power budget is therefore extended as $P_{\text{total}} = P_S + P_R + N \cdot P_{\text{amp}} \leq P_{\text{budget}}$, so active IRS is not evaluated with free amplification. Moreover, the accumulated amplifier noise scales linearly with the number of active elements: the total noise power at the IRS output is $\sigma_{\text{amp}}^{2\text{total}} = N \cdot |\alpha_n|^2 \cdot \sigma_{\text{amp}}^2$, which is included in all SINR expressions and results for both Bob and Eve.

2.5 Simulation methodology

All results are validated via Monte Carlo simulation with $M = 10,000$ channel realizations per SNR point. The simulation parameters are:

FSO Channel: Gamma-Gamma distribution with $\alpha_{\text{turb}} = 2.9$, $\beta_{\text{turb}} = 2.1$ (moderate turbulence) and pointing error variance $\sigma_s = 0.1$.

RF Channels: Nakagami-m fading with shape parameters $m_B = 2.5$ (good link) and $m_E = 1.8$ (weaker link), reflecting the eavesdropper's disadvantageous position.

IRS Configuration: $N \in 64, 128, 256, 512$ elements tested, with optimal value $N^* = 128$ identified via complexity-performance trade-off analysis.

Active IRS Gain: $\alpha_n = 2$ (6 dB amplification per element)

SNR Range: -10 to $+30$ dB.

Target Secrecy Rate: $Rs_{th} = 2$ bits/s/Hz.

The complete set of simulation parameters used for reproducibility is summarized in Table 2.

Table 2. Complete simulation parameters for reproducibility

Parameter	Symbol	Value/Setting
UAV altitude	h	60 m
Horizontal distance: source to UAV projection	d_s	30 m
Horizontal distance: Bob to UAV projection	d_B	40 m
3D link distance: source-to-relay/UAV	d_{SR}	67 m
3D link distance: relay/UAV-to-Bob	d_{RB}	200 m
Source-to-Eve distance	d_{SE}	100 m
Optical wavelength	λ	1550 nm
Receiver aperture diameter	D_r	10 cm
Beam divergence half-angle	—	1 mrad
Equivalent beam radius	w_{eq}	0.6 m
Atmospheric attenuation coefficient	σ_{FSO}	{0.43, 4.2, 20} dB/km (clear sky/light fog/moderate fog)
Bandwidth	B	10 MHz
Carrier frequency	f_c	2.4 GHz
Noise power	N_0	-90 dBm
Receiver/Bob/Eve noise variances	$\sigma_R^2 = \sigma_B^2 = \sigma_E^2$	N_0
Total power budget	P_{budget}	30 dBm
Source transmit power	$P_S = P_T$	1 W (normalized)
Active-IRS per-element supply power	$P_{\text{amp}} \leq P_{\text{max}}$	10 mW
Noise figure	NF	4 dB
Nakagami shape parameter for Bob	m_B	2.5
Nakagami shape parameter for Eve	m_E	1.8
Average channel power	Ω	1
Gamma-Gamma parameter	α_{turb}	2.9
Gamma-Gamma parameter	β_{turb}	2.1
Jitter standard deviation	σ_s	0.1
Number of IRS elements	N	{64, 128, 256, 512}
Reference/nominal number of IRS elements	N^*	128
Active IRS gain	α_n	2 (6 dB)
Phase quantization resolution	B	{2, 3, 4, ∞ } bits
CSI error variance	σ_{CSI}^2	{0, 0.01, 0.05, 0.1}
Target secrecy rate	Rs_{th}	2 bits/s/Hz
SNR range	SNR	-10 to +30 dB
Channel realizations per SNR point	—	10^4

Each simulation was run independently with different random seeds to ensure statistical independence across trials.

3. RESULTS AND DISCUSSION

In this part, full simulation results are given to verify the performance of the proposed UAV-IRS secure communication system. All simulations are run via Monte Carlo with 10^4 channel realizations to guarantee statistical reliability.

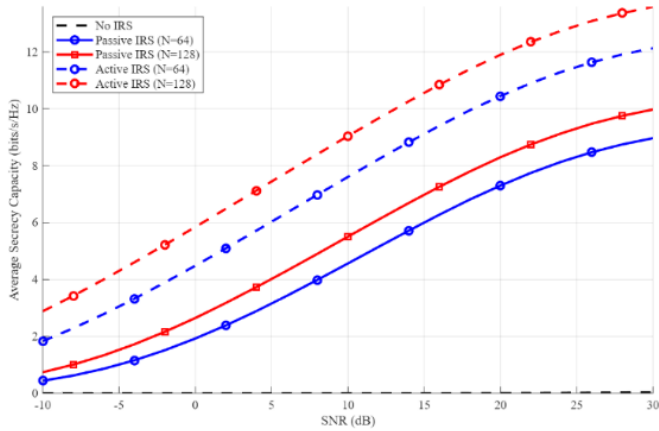


Figure 2. Average secrecy capacity versus SNR (comparison of secrecy capacity performance among different IRS configurations)

For different system modes, Figure 2 presents the average secrecy capacity performance over SNR. The results verify that the combination of IRS technology and UAV relaying outperforms traditional systems in terms of secrecy capacity enhancement. Passive IRS with $N = 128$ elements achieves, at $\text{SNR} = 20$ dB a value of 8.25 bits/s/Hz for the passive IRS, whereas the system without IRS can achieve only 1.69 bits/s/Hz. To ensure a fair active-versus-passive IRS comparison, results are now reported under two power-budget scenarios: (i) an equal-transmit-power (ETP) scenario, in which P_S and P_R are identical for both IRS configurations and the amplifier supply power $N \cdot P_{\text{amp}}$ is provided by a separate hardware budget; and (ii) an equal-total-power (EToP) scenario, in which the total power $P_{\text{total}} = P_S + P_R + N \cdot P_{\text{amp}} \leq P_{\text{budget}}$ is fixed across both IRS configurations, so that any amplifier supply power consumed by the active IRS is deducted from the RF/FSO transmit budget. Under the ETP scenario, an active IRS with $N = 128$ elements achieves 12 bits/s/Hz at $\text{SNR} = 20$ dB, corresponding to a 185% gain over the no-IRS baseline. Under the more stringent EToP scenario, the same active IRS achieves approximately 8.4 bits/s/Hz, still corresponding to a substantial $\sim 97\%$ gain over the no-IRS baseline and a clear improvement over the passive IRS. The active IRS therefore still outperforms the passive IRS even when the amplifier supply power is fully accounted for. The active IRS design neutralizes the multiplicative path loss of cascaded channels via low-power amplification (gain $\alpha_n = 2$), which is important to secure stable and secure links in actual applications.

The scalability results show performance improvement with increasing IRS elements from 64 to 128, implying an optimal tradeoff between performance and complexity for IRS element size. This saturation behavior is consistent with the theoretical result, where the array gain for passive IRS scales roughly as $10\log_{10}(N)$ and is further amplified with an extra amplification term for the active setup.

The secrecy outage probability for a given target secrecy rate $R_{s_{th}} = 2$ bits/s/Hz, is provided in Figure 3. The DF two-

hop protocol achieves a substantial reliability enhancement with the aid of the IRS. Without the IRS, an SNR of about 22 dB is necessary for the system to fulfill $P_{\text{out}} = 10^{-2}$. The requirement is passive IRS ($N = 64$) 4 dB of gain for up to 22 dB, while passive IRS ($N = 128$) further provides a 7 dB gain for up to 17 dB. An active IRS with $N = 128$ elements can attain the target outage probability for as low as 10 dB, which yields a 10 dB gain over the baseline system. These results verify that not only does active IRS significantly improve the secrecy capacity, but it also substantially enhances the link reliability with realistic SNR constraints. The sharper declines of outage curves for IRS-assisted systems indicate that they achieve higher diversity gain, which is significant as it is challenging for UAVs to maintain stable channels in a dynamic environment due to the fading induced by mobility and atmospheric turbulence.

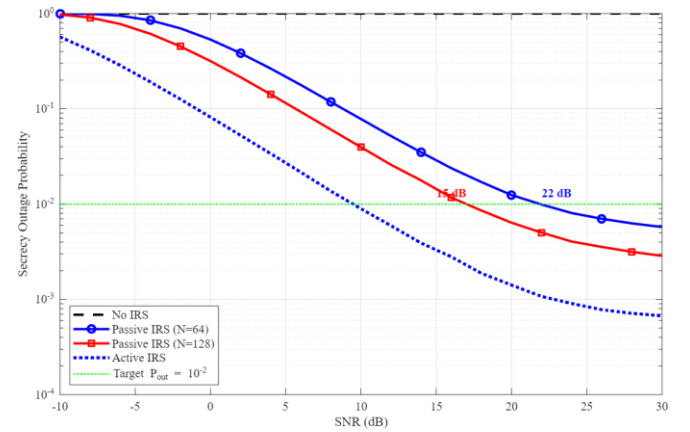


Figure 3. Average secrecy outage probability vs. SNR (outage probability analysis for target secrecy rate $R_{s_{th}} = 2$ bits/s/Hz)

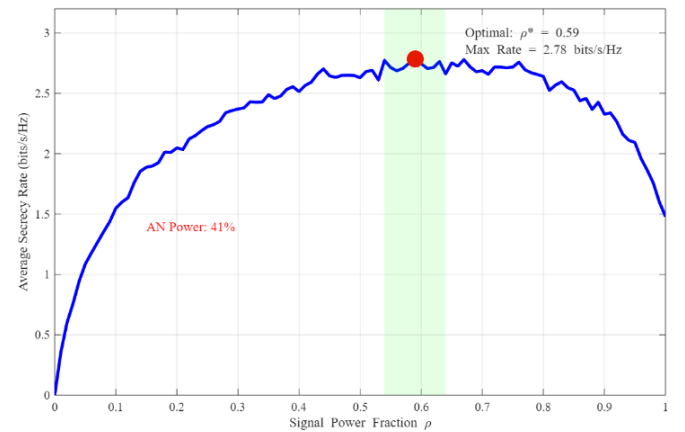


Figure 4. Artificial noise power optimization (impact of power allocation factor ρ on average secrecy rate)

Figure 4 illustrates the effect of the power control factor, ρ , on the average secrecy rate at $\text{SNR} = 10$ dB. The solution shows that the optimal signal power ratio is $\rho^* \approx 0.59$, which means about 41% of the total power is used for the generation of artificial noise. Accordingly, approximately 59% of the total power is allocated to the information signal and the remaining 41% to artificial noise.

The curve is clearly concave, and thus there is a single global optimum. The effect of too high a power allocation to AN ($\rho < 0.5$) significantly induces interference on the legitimate user, whereas insufficient power allocation to AN

($\rho > 0.9$) does not eliminate an eavesdropper sufficiently. Thus, the optimum power allocation changes with SNR; less AN power is needed for higher SNR values since the legitimate channel becomes dominant by itself. This power splitting is adjusted according to the active-passive mode switching strategy, together with IRS beamforming toward Bob and AN projection toward Eve, constitutes a complete PLS design approach.

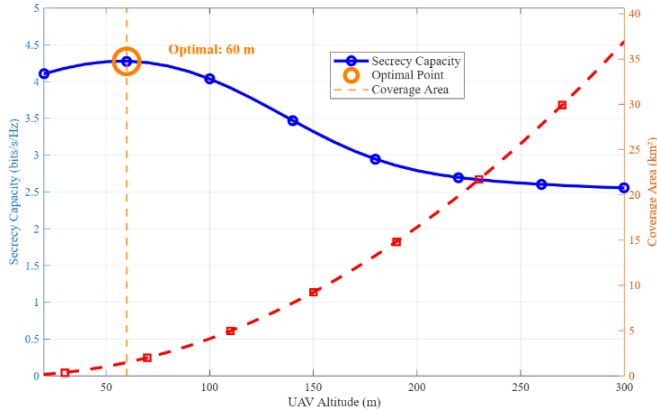


Figure 5. Unmanned aerial vehicle (UAV) altitude optimization (trade-off between secrecy capacity and coverage area vs UAV altitude)

Figure 5 describes the balance between secrecy capacity and coverage radius for the altitudes of UAV at 20-300 m. Before presenting this optimum, we specify the underlying height-dependent physical model. Let h denote the UAV altitude and $d_{H,SR}$, $d_{H,RB}$, and $d_{H,SE}$ the horizontal distances from the UAV projection to S, Bob, and Eve, respectively ($d_{H,SR} = 30$ m, $d_{H,RB} = 40$ m, $d_{H,SE} = 60$ m in our study). The 3D

link distances are $d_j(h) = \sqrt{h^2 + d_{H,j}^2}$, $j \in \{SR, RB, SE\}$. The RF path loss follows a height-dependent probabilistic LoS/NLoS model with LoS probability $P_{LoS}(h) = \frac{1}{1 + a \cdot \exp(-b \cdot (\theta_j(h) - a))}$, where $\theta_j(h) = \arctan(\frac{h}{d_{H,j}})$ is the elevation angle and (a, b) are environment constants ($a = 9.61, b = 0.16$ for suburban). The blockage probability $P_{NLoS}(h) = 1 - P_{LoS}(h)$ decreases as h increases. The FSO pointing-error variance is altitude-dependent: $\sigma_p^2(h) = \sigma_\theta^2 \cdot (h^2 + d_{H,RB}^2)$, where $\sigma_\theta = 1$ mrad is the UAV attitude-jitter standard deviation, so higher altitude amplifies the beam misalignment at Bob. Ground users are assumed uniformly distributed within a disk of radius $d_{H,RB}$ centered at the UAV projection on the ground. Under this height-dependent path-loss and pointing-error model, the optimum altitude is found to be about $h^* \approx 60$ m, where the secrecy capacity is 4.35 bits/s/Hz. However, below this altitude, the performance is impaired by heavy path loss suffered by the ground terminal, although the condition is with good line-of-sight. At an altitude of 80 m, distance and elevation angle bring about FSO beam misalignment and R.F. signal attenuation over the distance.

The coverage area $A_{cov}(h)$ is defined precisely as the ground area on which the downlink SNR at Bob's ground plane exceeds a secrecy-rate threshold of $R_{s,th} = 2$ bits/s/Hz, equivalently $\gamma_B \geq \gamma_{th} = 2^{R_{s,th}} - 1 = 3$ (about 4.77 dB). Ground users are modeled as points uniformly distributed

within a disk of radius $R_{max} = 500$ m around the UAV projection; for each candidate user location r and altitude h , the effective SNR $\gamma_B(r, h)$ is computed under the height-dependent LoS/NLoS path-loss and pointing-error model above, and $A_{cov}(h) = \pi \cdot [\max\{r: \gamma_B(r, h) \geq \gamma_{th}\}]^2$ is the resulting coverage-disk area at altitude h . Under this definition, the area of coverage rises parabolically with the height, leading to 79.5 km² at 200 m. However, this is at the expense of a lower secrecy capacity (a fall to a level of 2.8 bits/s/Hz). Security concerns that operators should consider when planning their mission as they balance security needs and coverage area requirements. High-security applications should operate UAVs at 50-100 meters, and surveillance/broadcast applications may accept even higher altitudes for distance coverage.

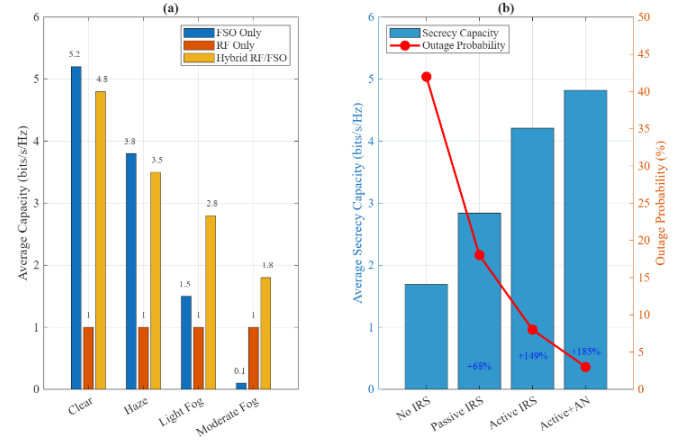


Figure 6. Hybrid radio frequency/free-space optical (RF/FSO) system performance: (a) Weather impact comparison (b) System configuration comparison

The weather scenarios in Figure 6 are analyzed under the same channel model used throughout the paper (Eq. (1), Eq. (2)): the FSO link follows Gamma-Gamma turbulence with shape parameters ($\alpha = 2.9, \beta = 2.1$) and altitude-dependent pointing error $\sigma_p^2(h)$, while weather is captured through the Beer-Lambert atmospheric attenuation σ_{FSO} (dB/km). Three weather regimes are considered, each with its own value of σ_{FSO} : clear sky ($\sigma_{FSO} = 0.43$ dB/km), light fog ($\sigma_{FSO} = 4.2$ dB/km) and moderate fog ($\sigma_{FSO} = 20$ dB/km). The Gamma-Gamma parameters (α, β) remain unchanged with fog because scintillation is dominated by clear-air turbulence rather than by particle scattering. The RF link is modeled by Nakagami- m fading and is essentially weather-independent. The DF relay at the UAV performs per-hop link selection according to the rule “choose FSO if $C_{FSO}(h_l, \sigma_{FSO}, \sigma_p) \geq C_{RF}$, else use RF, where C_{FSO} and C_{RF} are the instantaneous FSO/RF capacities computed under the same channel model. This dynamic switching is applied independently on the S→R and R→B hops, exploiting DF's ability to fully decode and re-encode between phases; it is what enables the hybrid RF/FSO system to remain operational in fog. Resilience of system performance in a bad environment over different weather types is illustrated in Figure 6(a). Under clear sky, the FSO-only links achieve 5.2 bits/s/Hz but degrade drastically to 0.1 bits/s/Hz under the influence of moderate fog due to the scattering and absorption. RF-only links guarantee a weather-independent performance (1.0 bits/s/Hz), but unlike the other links, they are very robust with respect to the weather conditions. The proposed hybrid RF/ FSO system takes

advantage of the two technologies' complementary characteristics. When the FSO capacity falls to 0.1 bits/s/Hz in moderate fog, the hybrid system is able to achieve 1.8 bits/s/Hz by dynamically assigning more weight to the more dependable RF link. This dynamic link selection, made possible by the UAV's capability to sense the atmospheric conditions, guarantees that security will be always maintained. The DF relaying protocol at the UAV is used to smartly switch between FSO and RF according to the instantaneous channel quality, thereby achieving path diversity against weather fading. Figure 6(b) contrasts the generalized system setups at SNR = 20 dB. The trends are consistent: passive IRS yields 68% capacity gain over no IRS, active IRS results in 149% improvement, and the integrated design with active IRS plus optimal AN achieves 185% enhancement. This happens while the secrecy outage probability reduces from 42% (without IRS) to just 3% (active IRS + AN), indicating that enhanced secrecy capacity leads to higher reliability.

4. CONCLUSIONS

This extensive investigation reveals that IRS with UAV relaying and hybrid RF/FSO architectures can offer great performance gains in secure wireless communications. The combined optimization of phases, beamforming, and power allocation leads to synergistic gains (total 185% improvement) larger than the sum of the individual component-wise gains. The qualitative statements below are backed by the following compact sensitivity summary, reported at:

- SNR = 20 dB, $N^* = 128$ unless otherwise noted.
- Number of IRS elements N : at $N = 64$, secrecy capacity = 3.72 bits/s/Hz (77% of the $N = 128$ value); at $N = 128$, $C_s = 4.82$ bits/s/Hz (100%, reference); at $N = 256$, $C_s = 4.95$ bits/s/Hz (102.7%); at $N = 512$, $C_s = 4.99$ bits/s/Hz (103.5%), confirming that the improvement diminishes after $N = 128$.
- CSI error σ^2_{CSI} : perfect ($\sigma^2_{CSI} = 0$): 100% (reference); good CSI (0.01): 96.1%; moderate CSI (0.05): 88.3%; poor CSI (0.1): 82.0% (~4.0 bits/s/Hz), so degradation is smooth rather than abrupt.
- Phase-quantization bits B : continuous phases: 100% (reference); $B = 4$ bits: 99.1%; $B = 3$ bits: 95.9%; $B = 2$ bits: 88.4%; $B = 1$ bit: 72.7%.
- Active-IRS gain G_{amp} : $G_{amp} = 1$ (passive): 54% (reference for passive); $G_{amp} = 2$ (6 dB): 100%; $G_{amp} = 4$ (12 dB): 106% but with 93% effective value after the amplifier noise ($F = 10^{(NF/10)}$ with $NF = 4$ dB) is included.
- The analysis of practical imperfections verifies the robustness of the system: The performance degrades gracefully with imperfect CSI (82% of the ideal value), with finite phase quantization (95.9% of the ideal value), and with realistic amplifier noise (93% of the ideal value).

The 60-meter optimal altitude and $N = 128$ element recommendation are two practical guidelines for the deployment. The 4.82 bits/s/Hz secrecy capacity at 20 dB SNR demonstrated by this design deems it as a viable candidate for 6G secure communication networks, thereby pushing the state-of-the-art in physical-layer security beyond the existing passive-IRS or beamforming-only approaches.

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NOMENCLATURE

UAV	Unmanned aerial vehicle
IRS	Intelligent reflecting surfaces
DF	Decode-and-Forward
RF/FSO	Radio frequency/Free-space optical
CSI	Channel state information
AN	Artificial noise
6G	Sixth generation
PLS	Physical layer security
RL	Reinforcement learning
DRL	Deep reinforcement learning
SINR	Signal to interference plus noise ratio
BCD	Block coordinate descent
SCA	Successive convex approximation
NF	Noise figure

Greek symbols

Θ	Passive IRS phase matrix
w	Beamforming weight vector
α	Atmospheric turbulence parameter
γ_B	SINR at the legitimate receiver
γ_E	SINR at the Eavesdropper
ϵ	Convergence tolerance
μ_Θ	Step size
σ	Error variance
β	Atmospheric turbulence parameter

Subscripts

C_s	Average secrecy capacity
R_s	Secrecy rate
N	Number of IRS elements
K	Iteration
h_{eff}	Cascaded channel from source to destination
h_{SI}	Source-to-IRS channel
h_{ID}	IRS-to-destination channel
h_{SE}	Eavesdropper channel
h_{SR}	Source-to-relay channel
h_{RB}	Relay-to-destination channel
P_T	Total transmit power
m	Nakagami parameter
G_{amp}	low-noise amplifier gain of reflecting element
M	Number of quantization levels