Performance analysis of RIS-Assisted Relay Systems

by

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Abstract

Reconfigurable intelligent surfaces (RISs) have been viewed as a vital physical-layer transmission technology for next-generation wireless systems. While the current state-of-art wireless standard is the fifth generation (5G), the RISs have emerged as a novel option to satisfy the rising demands for higher data rates, improved network coverage, increased reliability, and energy efficiency in wireless communications for sixth-generation (6G) and beyond. The need for greener communication technologies for wireless networks has laid the groundwork for many innovative wireless power transfer methods. Because radio-frequency (RF) signals can convey both information and energy simultaneously, there has been much research interest in designing novel technologies for simultaneous wireless information and power transmission (SWIPT) and energy harvesting (EH).
First, an RIS-assisted relay system model is proposed to improve the wireless system performance. By characterizing the optimal signal-to-noise ratio (SNR) attained through intelligent phase-shift controlling, the performance of the RIS-assisted relay system is investigated. Then, the performance of simultaneous wireless information and power transfer (SWIPT) is explored for the proposed RIS-assisted relay system. Also, the performance of linear EH models and non-linear EH models are compared via analytical and Monte-Carlo simulation results.

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  - Numerical results
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Chapter 1

Introduction

Motivation

- Spectral and energy efficiencies are two fundamental aspects of wireless communication system designs.
- **Reconfigurable intelligent surfaces (RISs)** have been identified as a key enabler for massive connectivity in the next-generation wireless systems standards.
    - Controllable environment: Allows to control phase-shifts of incident EM waves via passive reflective elements
    - Enhancing energy efficiency: Enables recycling of EM waves without generating additional signals via radio-frequency (RF) chains/amplifiers
    - High achievable rates: Enables to have constructive addition of EM waves at a desired destination
- **Relaying** can effectively reduce the end-to-end path-loss in terms of shorter-hop distances and amplify-and-forward (AF) or decode-and-forward (DF) operations at intermediate relays.

Motivation

- Radio-frequency (RF) signals can concurrently convey both information and energy.
- wireless nodes in the emerging Internet-of-Things (IoTs) many be able to perform data transmission, while supporting energy harvesting (EH) capabilities through the novel concept of **simultaneous wireless information and power transfer (SWIPT)**
- SWIPT technology relieve the energy consumption burden at the energy-constraint wireless devices.
- Prolonges the battery life of low-powered wireless nodes, which will be beneficial in powering billions of devices in the upcoming IoT era.

Reconfigurable intelligent surface (RIS)

- An **Reconfigurable intelligent surface (RIS)** having a large number of tiny passive reflectors can enable a controllable wireless propagation environment by introducing distinct delays to the reflected electromagnetic (EM) waves.
- These delays in turn result in controllable phase-shifts, which can be used to intelligently reconfigure propagation properties of EM waves through the wireless medium.
- This feature of RISs can be utilized to improve the signal-to-noise ratio (SNR) of an end-to-end communication between a transmitter and a receiver by enabling constructive additions of EM waves at a desired destination.

RIS in a single-hop communication scenario

- System model:
  - A single-antenna source (S)
  - A single-antenna destination (D)
  - An RIS having N reflectors
- Channel model:
  - h_m: the channel between S and the RIS
  - g_m: the channel between the mth RIS and D
- The polar-form of these channels

$$u = \beta_u e^{j\theta_u},$$

where $u \in \{h_m, g_m\}$ and $m \in \mathcal{M}$.

RIS-aided single hop set-up

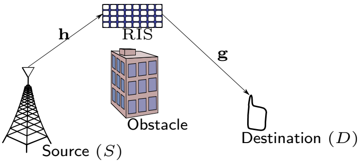


Figure: System and channel models for an RIS-aided Communication system.

- β_u denotes the envelop of u, and θ_u is the phase of u.
- β_u is assumed to be independent Rayleigh distributed as $f_{\beta_u}(x) = (x/\xi_u)\exp\left(-x^2/(2\xi_u)\right)$,

  where $\xi_u = \zeta_u/2$ is the Rayleigh parameter, and ζ_u accounts for the large-scale fading/path-loss of the channel u.

Signal Model

- If x is the signal transmitted by S, then the baseband signal received at D can be expressed as,

$$\begin{aligned} y_D &= \sqrt{P_t}\sum_{m=1}^{M} g_m \eta_m \mathrm{e}^{j\theta_{\eta_m}} h_m x + n_D \\ &= = \sqrt{P_t}\mathbf{g}^T \mathbf{\Theta} \mathbf{h} + n_D, \end{aligned}$$

where n_D is the additive white Gaussian noise (AWGN) at the destination such that $n_D \sim \mathcal{CN}(0, \sigma_D^2)$. Moreover, P_t is the transmit power at S, and η_m and θ_{η_m} are the reflection coefficient and phase-shifts introduced by the m_{th} reflecting element of RIS, respectively.
$\mathbf{h} = [h_1, \cdots, h_m, \cdots, h_M]^T$ is the channel vector between S and RIS, and $\mathbf{g} = [g_1, \cdots, g_m, \cdots, g_M]^T$ is the channel vector between RIS and D.
$\mathbf{\Theta} = \mathrm{diag}\left(\left(\eta_1 \mathrm{e}^{j\theta_{\eta_1}}, \cdots, \eta_m \mathrm{e}^{j\theta_{\eta_m}}, \cdots, \eta_M \mathrm{e}^{j\theta_{\eta_M}}\right)\right)$ is a diagonal matrix that captures the reflection properties of M reflecting elements of RIS.

Signal to noise ratio (SNR)

- The signal-to-noise ratio (SNR) at the destination can be written as

$$\gamma = \frac{P_t | \sum_{m=1}^{M} g_m \eta_m e^{j\theta_{\eta_m}} h_m . |^2}{\sigma_D^2}.$$

- Since the RIS provides an adjustable phase shift, we can co-phase the signals reflected by RIS setting $\theta_{\eta_m} = \theta_{h_m} + \theta_{g_m}$ to maximize the SNR. Upon optimizing the phase-shift matrix at the RIS, the optimal SNR is given by

$$\gamma = \bar{\gamma} \left(\sum_{m=1}^{M} \beta_{h_m} \eta_m \beta_{g_m} \right)^2,$$

where $\bar{\gamma} = P_t / \sigma_D^2$ is the average transmit SNR.

Energy harvesting protocol for single-hop networks

- In TS, the receiver structure changes between energy harvesting and information decoding modes on a regular basis.
  - $\alpha \in \{0, 1\}$ is the TS factor at the receivers for periodically switching between two separate modes.
- In PS, the receiver power is split into two portions by PS factor.
  - $\psi \in \{0, 1\}$ is the power switching factor.
- Hybrid Protocol is the combination of TS and PS protocol.
  - At $\alpha = 0$, Hybrid protocol $\rightarrow$ TS protocol.
  - At $\psi = 0$, Hybrid protocol $\rightarrow$ PS protocol.

RIS-aided single hop set-up

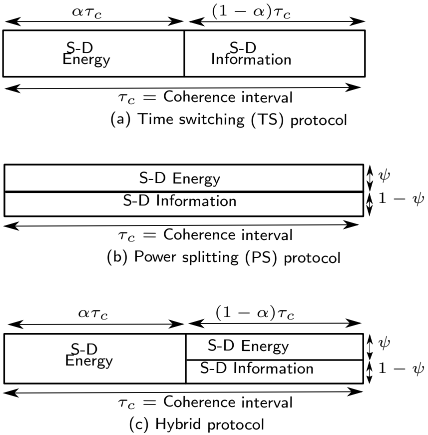


Figure: Single-hop energy harvesting protocol.

Chapter 2

Performance analysis of RIS-assisted Relay Networks

Motivation and contribution

- The fundamental performance metrics of the RIS-relay cascaded communication systems have not yet been investigated in the open literature.
- Our objective is to develop an analytical framework of an RIS-assisted relay system by deriving the performance bounds pertaining to the proposed RIS-relay cascaded system.
- First, the end-to-end optimal SNR is probabilistically characterized by tightly approximating it by a mathematically tractable counterpart by invoking the central limit theorem (CLT).
- Thereby, a tight upper bound for the cumulative distribution function (CDF) of this approximated optimal SNR is derived.
- By using this CDF, tight bounds/approximations for the average achievable rate, SNR/rate outage probability, and average SER are derived in closed-form.
- Then, the tightness of our performance bounds/approximations is validated through Monte-Carlo simulations.
- Finally, a set of insightful numerical results is presented to explore the performance gains of the proposed RIS-assisted relay system.

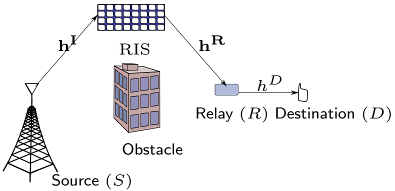
System and channel model

- System model:
  - A single-antenna source (S)
  - A single-antenna destination (D)
  - A single-antenna relay (R)
  - An RIS having N reflectors
- Channel model:
  - h_n^I: the channel between S and the nth reflector of the RIS
  - h_n^R: the channel between the nth reflector and R
  - h^D: the channel between R and D
- The polar-form of these channels

$$u = \beta_u e^{j\theta_u},$$

where $u \in \{h_n^I, h_n^R, h^D\}$ and $n \in \mathcal{N}$.

RIS-aided cell-free set-up



- β_u denotes the envelop of u, and θ_u is the phase of u.
- β_u is assumed to be independent Rayleigh distributed as

$$f_{\beta_u}(x) = (x/\xi_u) \exp\left(-x^2/(2\xi_u)\right),$$

where $\xi_u = \zeta_u/2$ is the Rayleigh parameter, and ζ_u accounts for the large-scale fading/path-loss of the channel u.

Signal model

- The signal transmitted by S reaches D through the RIS-R cascaded channel.
- The signal received at R during the first time-slot as

$$y_R = \sqrt{P}(\mathbf{h}^R)^T \mathbf{\Theta} \mathbf{h}^I x + w_R,$$

  - x: the transmit signal from S satisfying $\mathbb{E}\left[|x|^2\right] = 1$
  - P: the transmit power at S
  - $w_R \sim \mathcal{CN}\left(0, \sigma_{w_R}^2\right)$:: AWGN at R
  - $\mathbf{h}^I = [h_1^I, \cdots, h_n^I, \cdots, h_N^I]^T \in \mathbb{C}^{N\times 1}$
  - $(\mathbf{h}^R)^T = [h_1^R, \cdots, h_n^R, \cdots, h_N^R] \in \mathbb{C}^{1\times N}$
  - $\mathbf{\Theta} = \mathrm{diag}\left(\eta_1 \mathrm{e}^{j\theta_1}, \cdots, \eta_n \mathrm{e}^{j\theta_n}, \cdots, \eta_N \mathrm{e}^{j\theta_N}\right) \in \mathbb{C}^{N\times N}$: the reflective properties of the RIS, $\eta_n \mathrm{e}^{j\theta_n}$, represents the complex-valued reflection coefficient of the nth reflector of the RIS.
- By exploiting the properties of $\mathbf{\Theta}$, the rearranged received signal at R

$$y_R = \sqrt{P} \sum\nolimits_{n\in\mathcal{N}} h_n^R \eta_n \mathrm{e}^{j\theta_n} h_n^I x + w_R.$$

Signal model continued

- During the second time-slot, R first amplifies its received signal and then forwards it towards D.
- the signal received at D can be written as

$$y_D = Gh^D y_R + w_D = \sqrt{P} G h^D \sum_{n \in \mathcal{N}} h_n^R \eta_n \mathrm{e}^{j\theta_n} h_n^I x + G h^D w_R + w_D,$$

  where $w_D \sim \mathcal{CN}\left(0, \sigma_{w_D}^2\right)$ is an AWGN at D.
- G denotes the relay amplification factor, which is designed to constraint the instantaneous transmit power (P_R) at R

$$G = \sqrt{P_R \Big/ \left(P \left|\sum_{n \in \mathcal{N}} \eta_n \beta_{h_n^R} \beta_{h_n^I} \mathrm{e}^{j\phi_n}\right|^2 + \sigma_{w_R}^2\right)},$$

  where $\phi_n = \theta_n + \theta_{h_n^R} + \theta_{h_n^I}$.
- the received SNR at D as

$$\gamma = \frac{P \left|G h^D \sum_{n \in \mathcal{N}} h_n^R \eta_n \mathrm{e}^{j\theta_n} h_n^I\right|^2}{\left|G h^D\right|^2 \sigma_{w_R}^2 + \sigma_{w_D}^2}.$$

Signal model continued

- This SNR in terms of the channel phases

$$\gamma = \frac{P \left| G\beta_{h^D} \mathrm{e}^{j\theta_{h^D}} \sum_{n\in\mathcal{N}} \eta_n \beta_{h_n^R} \beta_{h_n^I} \mathrm{e}^{j\phi_n} \right|^2}{\left| G\beta_{h^D} \mathrm{e}^{j\theta_{h^D}} \right|^2 \sigma_{w_R}^2 + \sigma_{w_D}^2}.$$

- We maximize the received SNR at D by smartly adjusting the phase-shifts (θ_n) at each reflector to enable constructive addition of the signal terms inside the summation.

- The optimal choice of θ_n to maximize the received SNR

$$\theta_n^* = \underset{-\pi \le \theta_n \le \pi}{\operatorname{argmax}}\ \gamma = -\left(\theta_{h_n^R} + \theta_{h_n^I}\right), \quad \text{for} \quad n \in \mathcal{N}.$$

- The optimal SNR at D

$$\gamma^* = \frac{P\left(G^*\right)^2 \beta_{h^D}^2 \left(\sum_{n\in\mathcal{N}} \eta_n \beta_{h_n^R} \beta_{h_n^I}\right)^2}{\left(G^*\right)^2 \beta_{h^D}^2 \sigma_{w_R}^2 + \sigma_{w_D}^2} = \frac{\bar{\gamma}_R \left(\sum_{n\in\mathcal{N}} \eta_n \beta_{h_n^R} \beta_{h_n^I}\right)^2 \bar{\gamma}_D \beta_{h^D}^2}{\bar{\gamma}_R \left(\sum_{n\in\mathcal{N}} \eta_n \beta_{h_n^R} \beta_{h_n^I}\right)^2 + \bar{\gamma}_D \beta_{h^D}^2 + 1},$$

where $\bar{\gamma}_R = P_R/\sigma_{w_R}^2$ and $\bar{\gamma}_D = P/\sigma_{w_D}^2$.

Preliminary analysis

- The optimal received SNR is probabilistically characterized by deriving a tight approximate to its CDF.
- First, we define $Z = \sum_{n\in\mathcal{N}} \eta_n \beta_{h_n^R} \beta_{h_n^I}$.
- By using the fact that the envelops $\beta_{h_n^R}$ and $\beta_{h_n^I}$ are independent Rayleigh distributed random variables, Z is closely approximated by an one-sided Gaussian distributed random variable $(\tilde{Z})$ by invoking the CLT

$$f_Z(y) \approx f_{\tilde{Z}}(y) = \frac{\psi}{\sqrt{2\pi\sigma_Z^2}} \exp\left(\frac{-(y-\mu_Z)^2}{2\sigma_Z^2}\right), \text{ for } y \geq 0.$$

  - $\psi \triangleq 1/\mathcal{Q}\left(-\mu_Z/\sigma_Z\right)$: normalization factor, $\int_{-\infty}^{\infty} f_{\tilde{Y}}(x)dx = 1$
  - $\mu_Z = \sum_{n\in\mathcal{N}} \pi\eta_n \left(\xi_{h_n^R}\xi_{h_n^I}\right)^{1/2}/2$
  - $\sigma_Z^2 = \sum_{n\in\mathcal{N}} \eta_n^2 \xi_{h_n^R}\xi_{h_n^I}\left(16-\pi^2\right)/4$
- We define γ_R to be

$$\gamma_R = \bar{\gamma}_R Z^2 = \bar{\gamma}_R \left(\sum_{n\in\mathcal{N}} \eta_n \beta_{h_n^R} \beta_{h_n^I}\right)^2.$$

Preliminary analysis continued

- A tight approximation for the PDF of γ_R

$$f_{\gamma_R}(x) \approx \frac{\psi}{2\sqrt{\pi\sigma_R^2 x}} \exp\left(\frac{-(\sqrt{x}-\mu_R)^2}{2\sigma_R^2}\right), \text{ for } x \geq 0,$$

  where $\mu_R = \sqrt{\bar{\gamma}_R}\mu_Z$ and $\sigma_R^2 = \bar{\gamma}_R\sigma_Z^2$.
- An approximated CDF for γ_R

$$F_{\gamma_R}(x) \approx 1 - \psi\mathcal{Q}\left((\sqrt{x}-\mu_R)/\sigma_R\right), \text{ for } x \geq 0.$$

- Next, we define γ_D to be $\gamma_D = \bar{\gamma}_D\beta_{h^D}^2$. The CDF of γ_D

$$F_{\gamma_D}(x) = 1 - \exp\left(-x/\sigma_D^2\right), \text{ for } x \geq 0,$$

  where $\sigma_D^2 = \bar{\gamma}_D\zeta_{h^D}$.
- Since the exact derivation of the CDF of γ^* appears to be mathematically involved, we resort to an asymptotically exact upper bound

$$\gamma^* \approx \tilde{\gamma}^* = \min(\gamma_R, \gamma_D) = \min\left(\bar{\gamma}_R\left(\sum_{n\in\mathcal{N}}\eta_n\beta_{h_n^R}\beta_{h_n^I}\right)^2, \bar{\gamma}_D\beta_{h^D}^2\right).$$

Preliminary analysis continued

- By noticing that $\tilde{\gamma}^* = \min(\gamma_R, \gamma_D)$, the approximated CDF of γ^* (or the exact CDF of $\tilde{\gamma}^*$)

$$
\begin{aligned}
F_{\gamma^*}(y) &\approx F_{\tilde{\gamma}^*}(y) = 1 - (1 - F_{\gamma_R}(y))(1 - F_{\gamma_D}(y)) \\
&= 1 - \psi \mathcal{Q}\left((\sqrt{y} - \mu_R)/\sigma_R\right) \exp\left(-y/\sigma_D^2\right).
\end{aligned}
$$

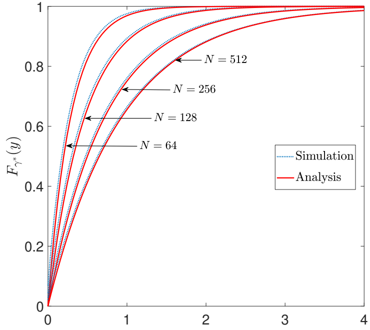


Figure: The CDF of SNR (γ^*) for $N \in \{64, 128, 256, 512\}$ and $\bar{\gamma} = 10$ dB.

Figure illustrates that our analytical CDF approximation is accurate for medium-to-large numbers of reflective elements (N) at the RIS. A relatively larger N is practically feasible and cost effective for RISs, and hence, our probabilistic characterization of the optimal SNR is useful in deriving performance bounds for the cascaded RIS-relay channels.

Average Achievable Rate

- The average achievable rate of the proposed system

$$\mathcal{R} = \mathbb{E}\left[\frac{1}{2}\log_2\left(1+\gamma^*\right)\right],$$

where the pre-log factor of $1/2$ is due to the fact that half-duplex relay mode requires two time-slots for end-to-end data transmission for the proposed system model.

- The exact derivation of $\mathcal{R}$ seems mathematically intractable, and hence, a tight upper bound by invoking the Jensen's inequality

$$\mathcal{R} \leq \mathcal{R}_{ub} = \frac{1}{2}\log_2\left(1+\mathbb{E}[\gamma^*]\right) \approx \frac{1}{2}\log_2\left(1+\mathbb{E}[\tilde{\gamma}^*]\right).$$

- The closed-form expressions of $\mathcal{R}_{ub}$

$$\mathcal{R}_{ub} = \frac{1}{2}\log_2\left(1+2\psi\sigma_D^2\mathcal{Q}\left(-\mu_R/\sigma_R\right)\right).$$

The SNR/rate outage probability

- For the proposed system, the probability that the instantaneous SNR (γ) falls below a threshold SNR (γ_{th}) is referred as the SNR outage probability.
- An approximation to this SNR outage probability

$$P_{out} = \Pr\left(\gamma^* \leq \gamma_{th}\right) \approx F_{\gamma^*}(\gamma_{th}).$$

- The rate outage probability can also be readily obtained as

$$\begin{aligned} P_{out} &= \Pr\left(\mathcal{R}' \leq \mathcal{R}_{th}\right) = \Pr\left(\gamma^* \leq 2^{2\mathcal{R}_{th}} - 1\right) \\ &\approx F_{\gamma^*}(2^{2\mathcal{R}_{th}} - 1), \end{aligned}$$

where $\mathcal{R}' = \frac{1}{2}\log_2\left(1 + \gamma^*\right)$ is the achievable rate, and $\mathcal{R}_{th}$ denotes a threshold rate.

The average symbol error rate (SER)

- The average SER of the proposed systems is defined as the expectation of conditional error probability $(P_{e|\gamma^*})$ over the probability distribution of γ^*.
- $P_{e|\gamma^*}$ is given for a broad range of coherent modulation schemes by $P_{e|\gamma^*} = \omega\mathcal{Q}\left(\sqrt{\vartheta\gamma^*}\right)$, where the modulation scheme determines the values of fixed parameters ω and ϑ.
- Thus, the average SER: $\bar{P}_e = \mathbb{E}\left[\omega\mathcal{Q}\left(\sqrt{\vartheta\gamma^*}\right)\right]$.
- A tight approximation for $\bar{P}_e$ can be given as

$$\bar{P}_e \approx \frac{\omega}{2} - \frac{\omega\sqrt{\vartheta}}{2\sqrt{2\pi}}\int_0^\infty x^{-1/2}\exp\left(-\frac{\vartheta x}{2}\right)\bar{F}_{\tilde{\gamma}^*}(x)dx,$$

where $\bar{F}_{\tilde{\gamma}^*}(x) = 1 - F_{\tilde{\gamma}^*}(x)$ is the CCDF of γ^*.

- The closed-form solution to $\bar{P}_e$

$$\begin{aligned}\bar{P}_e \;&\approx\; \frac{\omega}{2} + 2\lambda\mathcal{Q}\left(-\mu_R/\sigma_R\right) + \frac{\lambda}{\pi\sqrt{a\sigma_R^2}}\mathrm{e}^{\frac{-\mu_R^2}{2\sigma_R^2}\left(1-\frac{1}{2a\rho\sigma_R^2}\right)} \\ &\times\sum\nolimits_{i\in C_\infty}(-2)^i\left(\frac{\sqrt{a}\rho^2\sigma_R^2}{\mu_R}\right)^{i+1}\Gamma\left(\frac{i+1}{2}, \frac{\mu_R^2}{2a\rho\sigma_R^4}\right).\end{aligned}$$

Effects of Phase Quantization

- In practice, an RIS reflecting element only uses a set of discrete phase-shifts due to the associated hardware limitations.
- We investigate the impact of phase-shift quantization assuming that a limited number of discrete phase-shift is available for selection at the nth RIS element as $\hat{\theta}_n^* = \pi\hat{q}/2^{b-1}$,
  - b is the number of quantization bits
  - $\hat{q} = \underset{q\in\{0,\pm 1,\cdots,\pm 2^{b-1}\}}{\text{argmin}} |\theta_n^* - \pi q/2^{b-1}|$
  - $\hat{\theta}_n^*$ is the optimal phase-shift
- The difference between unquantized and quantized phase-shift is defined as the phase quantization error: $\epsilon_n = \theta_n^* - \hat{\theta}_n^*$.
- When the number of quantization levels increases, ϵ_n converges to a uniform distribution as $\epsilon_n \sim \mathcal{U}[-\pi/2^b, \pi/2^b)$.
- The optimal SNR with discrete phase-shift

$$\hat{\gamma}^* = \frac{\bar{\gamma}_R \bar{\gamma}_D \beta_{h^D}^2 \left(\sum_{n\in\mathcal{N}} \eta_n \beta_{h_n^R} \beta_{h_n^I} \mathrm{e}^{j\epsilon_n}\right)^2}{\bar{\gamma}_R \left(\sum_{n\in\mathcal{N}} \eta_n \beta_{h_n^R} \beta_{h_n^I} \mathrm{e}^{j\epsilon_n}\right)^2 + \bar{\gamma}_D \beta_{h^D}^2 + 1}.$$

Simulation: Outage probability

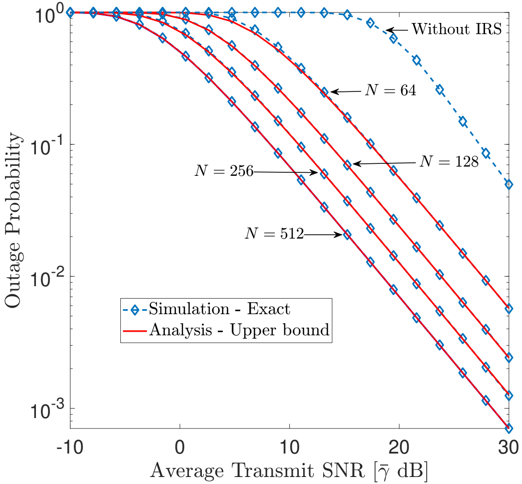


Figure: The outage probability for $N \in \{64, 128, 256, 512\}$ and the threshold SNR is $\gamma_{th} = 0$ dB. .

Simulation: Average achievable rate

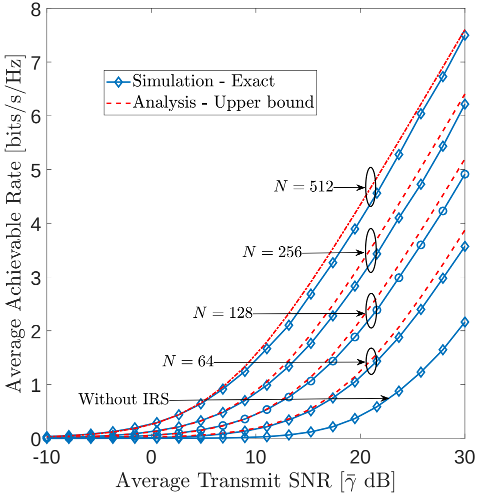


Figure: The average achievable rate for $N \in \{64, 128, 256, 512\}$.

Simulation: Average BER

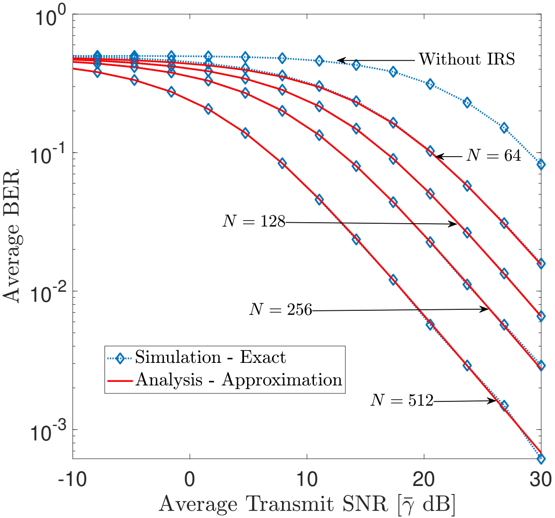


Figure: The average BER of BPSK for $N \in \{64, 128, 256, 512\}$, $\omega = 1$, and $\vartheta = 2$.

Simulation: Phase-shift quantization

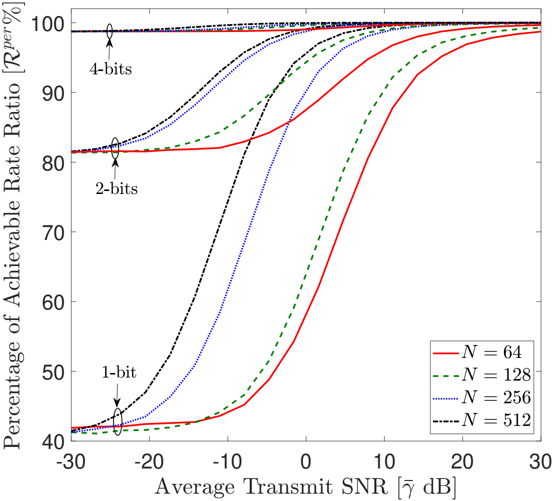


Figure: The effect of phase shift quantization on the average achievable rate for $N \in \{64, 128, 256, 512\}$.

Chapter 3

Energy Harvesting in RIS-Assisted Relay Networks

Motivation and Contribution

- Energy harvesting aspects and **Simultaneous Wireless Information and Power Transfer (SWIPT)** for RIS-relay cascaded communication systems have not been studied in the previous studies
- We aim to present an analytical approach to characterize the performance bounds of SWIPT technology for the RIS-relay cascaded systems adopting hybrid protocol.
- First, the optimal SNR expression is derived by optimizing the phase-shifts at the reflecting coefficients of the RIS.
- Then, we perform statistical characterization of the end-to-end SNR and derive tight bounds in closed-form for the average achievable rate and harvested energy for the hybrid SWIPT protocol.
- Moreover, we derive the achievable rate-energy trade-off, which is fundamental for SWIPT-enabled systems.
- Finally, the performance degradation of the proposed system model due to practical impairments is investigated by capturing the discrete phase-shift adjustments at the RIS.

System and channel model

- System model:
  - A single-antenna source (S)
  - A single-antenna destination (D) equipped with SWIPT receiver
  - A single-antenna relay (R)
  - An RIS having N reflectors
- Channel model:
  - h_n^I: the channel between S and the nth reflector of the RIS
  - h_n^R: the channel between the nth reflector and R
  - h^D: the channel between R and D

RIS-aided cell-free set-up

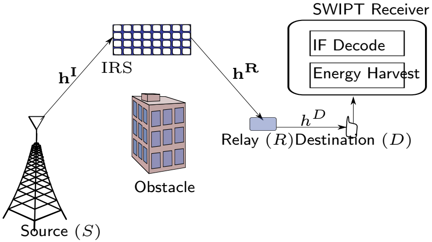


Figure: System model-RIS-assisted relay network with SWIPT at the receiver.

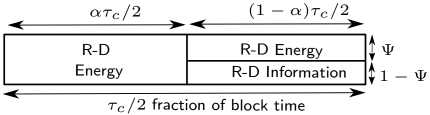


Figure: Time frame for $\tau_c/2$ fraction of coherence interval.

Signal model

- The half-duplex relay mode requires $\tau_c/2$ fraction of channel coherence time to receive the signal for the proposed system model.
- The signal received at R during this fraction of channel coherence time can be expressed as

$$y_R = \sqrt{P}(\mathbf{h}^R)^T \mathbf{\Theta} \mathbf{h}^I x + w_R,$$

  - x: the transmit signal from S satisfying $\mathbb{E}\left[|x|^2\right] = 1$
  - P: the transmit power at S
  - $w_R \sim \mathcal{CN}\left(0, \sigma_{w_R}^2\right)$:: AWGN at R
  - $\mathbf{h}^I = [h_1^I, \cdots, h_n^I, \cdots, h_N^I]^T \in \mathbb{C}^{N\times 1}$
  - $(\mathbf{h}^R)^T = [h_1^R, \cdots, h_n^R, \cdots, h_N^R] \in \mathbb{C}^{1\times N}$
  - $\mathbf{\Theta} = \mathrm{diag}\left(\eta_1 \mathrm{e}^{j\theta_1}, \cdots, \eta_n \mathrm{e}^{j\theta_n}, \cdots, \eta_N \mathrm{e}^{j\theta_N}\right) \in \mathbb{C}^{N\times N}$: the reflective properties of the RIS, $\eta_n \mathrm{e}^{j\theta_n}$, represents the complex-valued reflection coefficient of the nth reflector of the RIS.
- By exploiting the properties of $\mathbf{\Theta}$, the rearranged received signal at R

$$y_R = \sqrt{P} \sum\nolimits_{n\in\mathcal{N}} h_n^R \eta_n \mathrm{e}^{j\theta_n} h_n^I x + w_R.$$

Signal model continued

- During the next $\tau_c/2$ fraction of block time, R first amplifies its received signal and then forwards it towards D.
- the signal received at D can be written as

$$y_D = Gh^D y_R + w_D = \sqrt{P} G h^D \sum_{n \in \mathcal{N}} h_n^R \eta_n \mathrm{e}^{j\theta_n} h_n^I x + G h^D w_R + w_D,$$

where $w_D \sim \mathcal{CN}\left(0, \sigma_{w_D}^2\right)$ is an AWGN at D.
- G denotes the relay amplification factor, which is designed to constraint the instantaneous transmit power (P_R) at R. If we use fixed gain relaying using only statistical channel information, then G can be expressed as

$$G = \sqrt{P_R \Big/ \left(\mathbb{E}\left[\left|\sum_{n \in \mathcal{N}} \eta_n \beta_{h_n^R} \beta_{h_n^I} \mathrm{e}^{j\phi_n}\right|^2\right] P + \sigma_{w_R}^2\right)},$$

where $\phi_n = \theta_n + \theta_{h_n^R} + \theta_{h_n^I}$. Here G is constant average relay gain.

Energy Harvesting for linear model

- After S-R transmission during $\tau_c/2$ fraction of coherence time following the relay amplification process, the next $\alpha\tau_c/2$ fraction of the coherence time is utilized only for energy harvesting from the received signal.
- In the remaining $(1-\alpha)\tau_c/2$ fraction of coherence time, a Ψ portion of the power from the received signal is used for energy harvesting. Thus, for the linear model, the harvested energy can be expressed as

$$\begin{aligned} E_{LM} &= \Lambda P \left| G\beta_{h^D} \mathrm{e}^{j\theta_{h^D}} \sum\nolimits_{n\in\mathcal{N}} \beta_{h_n^R}\eta_n \mathrm{e}^{j\phi_n}\beta_{h_n^I}\right|^2 \alpha\tau_c/2 \\ &\quad + \Lambda\Psi P \left| G\beta_{h^D} \mathrm{e}^{j\theta_{h^D}} \sum\nolimits_{n\in\mathcal{N}} \beta_{h_n^R}\eta_n \mathrm{e}^{j\phi_n}\beta_{h_n^I}\right|^2 (1-\alpha)\tau_c/2, \end{aligned}$$

  where Λ is the RF to direct current conversion efficiency.
- We maximize the harvested energy at D using by smartly adjusting the phase-shifts (θ_n) at each reflector to allow constructive addition of the signal terms inside the summation term of above equation.

Energy Harvesting for linear model

- The optimal choice of θ_n to maximize the harvested energy at D is given by

$$\theta_n^* = \underset{-\pi \leq \theta_n \leq \pi}{\operatorname{argmax}} \ \gamma = -\left(\theta_{h_n^R} + \theta_{h_n^I}\right), \quad \text{for} \quad n \in \mathcal{N}.$$

- Thus, the optimal relay gain is given by

$$\begin{aligned} G^* &= \sqrt{P_R \Big/ \left(\mathbb{E}\left[\left|\sum\nolimits_{n\in\mathcal{N}} \eta_n \beta_{h_n^R} \beta_{h_n^I}\right|^2\right] P + \sigma_{w_R}^2\right)} \\ &= \sqrt{P_R / \left(\mathbb{E}[V^2]\, P + \sigma_{w_R}^2\right)}, \end{aligned}$$

where, $V = \sum_{n\in\mathcal{N}} \eta_n \beta_{h_n^R} \beta_{h_n^I}$, and $\mathbb{E}\left[V^2\right]$ is derived as

$$\mathbb{E}\left[V^2\right] = \left(1 - \pi^2/16\right) \sum_{n\in\mathcal{N}} \delta_n^2 + \pi^2/16 \left(\sum_{n\in\mathcal{N}} \delta_n\right)$$

Effects of Phase Quantization

- In order to maximize the harvested energy, we assume that each RIS element provides a continuous phase-shift.
- However in practice, an RIS element only uses a set of discrete phase-shifts due to the hardware limitations associated with it.
- Thus, we investigate the impact of phase-shift quantization assuming that a limited number of discrete phase-shift is available for selection at the nth RIS element as $\hat{\theta}_n^* = \pi\hat{q}/2^{b-1}$
  - b is the number of quantization bits
  - $\hat{q} = \underset{q\in\{0,\pm1,\cdots,\pm2^{b-1}\}}{\text{argmin}} |\theta_n^* - \pi q/2^{b-1}|$
  - $\hat{\theta}_n^*$ is the optimal phase-shift
- The difference between unquantized and quantized phase-shift is defined as the phase quantization error: $\epsilon_n = \theta_n^* - \hat{\theta}_n^*$.
- When the number of quantization levels increases, ϵ_n converges to a uniform distribution as $\epsilon_n \sim \mathcal{U}[-\pi/2^b, \pi/2^b)$.

Harvested Energy with discrete quantization

- Thus, the optimal relay gain with discrete phase-shift can be given as

$$\begin{aligned}\hat{G}^* &= \sqrt{P_R \Big/ \left(\mathbb{E}\left[\left|\sum_{n\in\mathcal{N}} \eta_n \beta_{h_n^R} \beta_{h_n^I} \mathrm{e}^{j\epsilon_n}\right|^2\right] P + \sigma_{w_R}^2\right)} \\ &= \sqrt{P_R / \left(\left(\mathbb{E}[W_R^2] + \mathbb{E}[W_I^2]\right) P + \sigma_{w_R}^2\right)},\end{aligned}$$

where $W_R = \sum_{n\in\mathcal{N}} \eta_n \beta_{h_n^R} \beta_{h_n^I} \cos\epsilon_n$, and $W_I = \sum_{n\in\mathcal{N}} \eta_n \beta_{h_n^R} \beta_{h_n^I} \sin\epsilon_n$.

- Thus, the optimal harvested energy with discrete phase-shift can be given as

$$\begin{aligned}E_{LM}^{opt} &= \Lambda P \left| G\beta_{h^D} \mathrm{e}^{j\theta_{h^D}} \sum_{n\in\mathcal{N}} \beta_{h_n^R} \eta_n \mathrm{e}^{j\phi_n} \beta_{h_n^I} \mathrm{e}^{j\epsilon_n} \right|^2 \alpha\tau_c/2 \\ &\quad + \Lambda\Psi P \left| G\beta_{h^D} \mathrm{e}^{j\theta_{h^D}} \sum_{n\in\mathcal{N}} \beta_{h_n^R} \eta_n \mathrm{e}^{j\phi_n} \beta_{h_n^I} \mathrm{e}^{j\epsilon_n} \right|^2 (1-\alpha)\tau_c/2 \\ &= [\alpha + (1-\alpha)\Psi]\Lambda \frac{\tau_c}{2} P |\hat{G}^*|^2 |\beta_{h^D}|^2 [W_R^2 + W_I^2].\end{aligned}$$

Average Harvested Energy

- The average optimal harvested energy using the linear model via mathematical and probabilistic derivation would be

$$
\begin{aligned}
\overline{E_{LM}^{opt}} &= \mathbb{E}\left[E_{LM}^{opt}\right] \\
&= [\alpha + (1-\alpha)\Psi]\Lambda \frac{\tau_c}{2} P(\hat{G}^*)^2 \zeta_{h_D} \\
&\times \left[\left(1 - \frac{\pi^2 \sin^2(\tau)}{16\tau^2}\right) \sum_{n \in \mathcal{N}} \delta_n^2 + \frac{\pi^2 \sin^2(\tau)}{16\tau^2} \left(\sum_{n \in \mathcal{N}} \delta_n\right)^2\right],
\end{aligned}
$$

where $\delta_n^2 = \left(\eta_n^2 \zeta_{h_n^I} \zeta_{h_n^R}\right)^2$.

Harvested Energy via non-linear model

- The harvested energy via non-linear model can be expressed as

$$H(P_D) = \left[\frac{K}{1-\omega}\left(\frac{1}{1+\mathrm{e}^{-a(P_D-b)}} - \omega\right)\right]^+,$$

  - $[Y]^+ = \max(0, Y)$
  - $\omega = 1/(1+\mathrm{e}^{ab})$

  Here P_D is the incident power.

- The non-linear effects of energy harvesting circuits are captured by parameters a, b, and K.
- For practical energy harvesters, these parameters have values: $a = 47.083 \times 10^{-3}$, $b = 2.9\mu W$ $K = 9.079\mu W$.
- We can quantify the total harvested energy for our system as

$$E_{NL} = \alpha\tau_c H(P_D) + (1-\alpha)\tau H(\Psi P_D).$$

Received Power via non-linear model

- The power received by the destination antenna in can be expressed as

$$P_D = P \left| G \beta_{h^D} \sum_{n \in \mathcal{N}} \beta_{h_n^R} \beta_{h_n^I} \eta_n \mathrm{e}^{j\epsilon_n} \right|^2$$

- As the function $H(P_D)$ is non-decreasing, optimizing P_D would maximize the energy harvested by the non-linear model.
- Using optimizing procedure for the phase-shifts at the RISs, we obtain the optimum P_D as

$$\begin{aligned} P_D^{opt} &= P \left| G^* \beta_{h^D} \sum_{n \in \mathcal{N}} \beta_{h_n^R} \beta_{h_n^I} \eta_n \mathrm{e}^{j\epsilon_n} \right|^2 \\ &= P \left[\left| G^* \beta_{h^D} \right|^2 \left(W_R^2 + W_I^2 \right) \right] \end{aligned}$$

Optimal received Energy

- Then, we can obtain the optimal harvested energy for non-linear model by invoking Jensen's inequality as

$$\overline{E_{NL}^{opt}} = \mathbb{E}\left[E_{NL}^{opt}\right] \leq \frac{\tau_c}{2} H\left(\overline{P_D^{opt}}\right) + (1-\alpha)\frac{\tau_c}{2} H\left(\Psi \overline{P_D^{opt}}\right),$$

- where the average optimal received power can be found by taking expectation in (1) as

$$\overline{P_D^{opt}} = \mathbb{E}\left[P_D^{opt}\right] = P|\hat{G}^*|^2 \mathbb{E}\left[|\beta_{h^D}|^2\right] \left(\mathbb{E}\left[W_R^2\right] + \mathbb{E}\left[W_I^2\right]\right).$$

- Mathematical manipulation gives us

$$\overline{P_D^{opt}} = P(\hat{G}^*)^2 \zeta_{h_D} \left[\left(1 - \frac{\pi^2 \sin^2(\tau)}{16\tau^2}\right) \sum_{n\in\mathcal{N}} \delta_n^2 + \frac{\pi^2 \sin^2(\tau)}{16\tau^2} \left(\sum_{n\in\mathcal{N}} \delta_n\right)^2\right].$$

Optimal received SNR

- During the $(1-\alpha)\tau_c/2$ fraction of the channel coherence time, a $1-\Psi$ portion of the received signal at the R is transmitted through the decoder.
- The SNR of the received signal at D can be derived as

$$\gamma = (1-\Psi)\frac{P\left|G\beta_{h^D}\mathrm{e}^{j\theta_{h^D}}\sum_{n\in\mathcal{N}}\eta_n\beta_{h_n^R}\beta_{h_n^I}\mathrm{e}^{j\phi_n}\right|^2}{\left|G\beta_{h^D}\mathrm{e}^{j\theta_{h^D}}\right|^2\sigma_{w_R}^2+\sigma_{w_D}^2}.$$

- We maximize the received SNR at D by smartly adjusting the phase-shifts (θ_n) at each reflector to enable constructive addition of the signal terms inside the summation.

$$\begin{aligned}\gamma^* &= (1-\Psi)\frac{P\left(G^*\right)^2\beta_{h^D}^2\left(\sum_{n\in\mathcal{N}}\eta_n\beta_{h_n^R}\beta_{h_n^I}\right)^2}{\left(G^*\right)^2\beta_{h^D}^2\sigma_{w_R}^2+\sigma_{w_D}^2} \\ &= (1-\Psi)\frac{\bar{\gamma}_R\bar{\gamma}_D\beta_{h^D}^2\left(\sum_{n\in\mathcal{N}}\eta_n\beta_{h_n^R}\beta_{h_n^I}\right)^2}{\bar{\gamma}_R\left(\sum_{n\in\mathcal{N}}\eta_n\beta_{h_n^R}\beta_{h_n^I}\right)^2+\bar{\gamma}_D\beta_{h^D}^2+1}, \\ &= (1-\Psi)\frac{\bar{\gamma}_R\bar{\gamma}_D\beta_{h^D}^2\left(W_R^2+W_I^2\right)}{\bar{\gamma}_R\left(W_R^2+W_I^2\right)+\bar{\gamma}_D\beta_{h^D}^2+1}.\end{aligned}$$

Preliminary analysis

- A tight approximation for the PDF of γ_R

$$f_{\gamma_R}(x) \approx \frac{\psi}{2\sqrt{\pi\sigma_R^2 x}} \exp\left(\frac{-(\sqrt{x}-\mu_R)^2}{2\sigma_R^2}\right), \text{ for } x \geq 0,$$

  where $\mu_R = \sqrt{\bar{\gamma}_R}\mu_Z$ and $\sigma_R^2 = \bar{\gamma}_R \sigma_Z^2$.
- An approximated CDF for γ_R

$$F_{\gamma_R}(x) \approx 1 - \psi \mathcal{Q}\left((\sqrt{x}-\mu_R)/\sigma_R\right), \text{ for } x \geq 0.$$

- Next, we define γ_D to be $\gamma_D = \bar{\gamma}_D \beta_{h^D}^2$. The CDF of γ_D

$$F_{\gamma_D}(x) = 1 - \exp\left(-x/\sigma_D^2\right), \text{ for } x \geq 0,$$

  where $\sigma_D^2 = \bar{\gamma}_D \zeta_{h^D}$.
- Since the exact derivation of the CDF of γ^* appears to be mathematically involved, we resort to an asymptotically exact upper bound

$$\gamma^* \approx \tilde{\gamma}^* = \min\left(\bar{\gamma}_R \left(\sum\nolimits_{n\in\mathcal{N}} \eta_n \beta_{h_n^R} \beta_{h_n^I}\right)^2, \bar{\gamma}_D \beta_{h^D}^2\right).$$

The CDF of received SNR

- since the exact derivation of the CDF of γ^* appears to be mathematically involved, we resort to an asymptotically exact upper bound as

$$\gamma^* \approx \tilde{\gamma}^* = (1-\Psi) \times \min\left(\bar{\gamma}_R \left(\sum\nolimits_{n\in\mathcal{N}} \eta_n \beta_{h_n^R} \beta_{h_n^I}\right)^2, \bar{\gamma}_D \beta_{h^D}^2\right).$$

- Using transformation of random variable technique, we obtain the SNR of $\tilde{\gamma}^* = (1-\Psi)\min(\gamma_R, \gamma_D)$ as

$$\begin{aligned} F_{\gamma^*}(y) \approx F_{\tilde{\gamma}^*}(y) &= F_X\left(\frac{y}{1-\Psi}\right) \\ &= 1-\psi\mathcal{Q}\left[\left(\sqrt{\frac{y}{1-\Psi}}-\mu_R\right)/\sigma_R\right] \mathrm{e}-\frac{y}{(1-\Psi)\sigma_D^2}. \end{aligned}$$

Average Achievable Rate

- The average achievable rate can be defined as follows:

$$\mathcal{R} = \left(\frac{1-\alpha}{2}\right)\mathbb{E}[\log_2\left(1+\gamma^*\right)],$$

where the pre-log factor of $(1-\alpha)/2$ captures the effective portion of coherence time for end-to-end data transmission for the proposed system model.

- The exact derivation of $\mathcal{R}$ appears mathematically intractable, and therefore, we again resort to a tight upper bound by using the Jensen's inequality as

$$\mathcal{R} \leq \mathcal{R}_{ub} = \left(\frac{1-\alpha}{2}\right)\log_2\left(1+\mathbb{E}[\gamma^*]\right) \approx \left(\frac{1-\alpha}{2}\right)\log_2\left(1+\mathbb{E}[\tilde{\gamma}^*]\right).$$

- Then, by evaluating the expectation term, an achievable rate upper bound is computed as

$$\mathcal{R}_{ub} = \left(\frac{1-\alpha}{2}\right)\log_2\left(1+2(1-\Psi)\psi\sigma_D^2\mathcal{Q}\left(-\mu_R/\sigma_R\right)\right).$$

Energy-Rate Trade-offs

- A tight upper bound for the optimal energy-rate trade-off of the TS protocol is determined for the linear energy harvesting model by allowing $\Psi = 0$, then solving for α, and finally replacing it in rate expression as

$$\mathcal{R}_{ub}^{TS} = \left[1 - \frac{\overline{E_{LM}^{opt}}/2}{\Lambda \frac{\tau_c}{2} P(\hat{G}^*)^2 \zeta_{h_D} \left[\left(1 - \frac{\pi^2 \sin^2(\tau)}{16\tau^2}\right) \sum_{n\in\mathcal{N}} \delta_n^2 + \frac{\pi^2 \sin^2(\tau)}{16\tau^2} \left(\sum_{n\in\mathcal{N}} \delta_n\right)^2\right]} \times \log_2\left(1 + 2\psi\sigma_D^2 \mathcal{Q}\left(-\frac{\mu_R}{\sigma_R}\right)\right),$$

- Similarly, a tight upper bound optimal energy-rate trade-off for the PS protocol with a linear energy harvesting model may be achieved by setting $\alpha = 0$ to find Ψ and finally inserting it into the rate expression as

$$\mathcal{R}_{ub}^{PS} = \frac{1}{2}\log_2\left(1 + 2\psi\sigma_D^2 \mathcal{Q}\left(\frac{-\mu_R}{\sigma_R}\right)\left(1 - \frac{\overline{E_{LM}^{opt}}/\left(\Lambda \frac{\tau_c}{2} P(\hat{G}^*)^2 \zeta_{h_D}\right)}{\left(1 - \frac{\pi^2 \sin^2(\tau)}{16\tau^2}\right)\sum_{n=0}^{N} \delta_n^2 + \frac{\pi^2 \sin^2(\tau)}{16\tau^2} \left(\sum_{n=0}^{N} \delta_n\right.}\right.\right.$$

Energy-Rate Trade-offs for non-linear models

- Similarly, a tight upper bound optimal energy-rate trade-off for non-linear harvesting model can also be found from similar mathematical procedures as

$$\left(\mathcal{R}_{ub}^{TS}\right)_{NL} = \left[1 - \frac{\overline{E_{LM}^{opt}}}{2\tau_c H\left(\overline{P_D^{opt}}\right)}\right],$$

$$\left(\mathcal{R}_{ub}^{PS}\right)_{NL} = \frac{1}{2}\log_2\left(1 + 2\psi\sigma_D^2\mathcal{Q}\left(-\frac{\mu_R}{\sigma_R}\right)\left(1 - \ln\left[\left(\frac{\mathrm{e}^{ab}\overline{E_{NL}^{opt}} + K\tau_c}{K\tau_c - \overline{E_{NL}^{opt}}}\right)\Big/ a\overline{P_D^{opt}}\right]\right.\right.$$

Continuous phase-shifts

- When $b \to \infty$, the phase-shift error ($\epsilon_n \to 0$). Therefore, we can derive the average harvested energy for the linear model by setting $\tau = 0$ in (1) as

$$\begin{aligned}\overline{E_{LM}^{opt}} &= \mathbb{E}\left[E_{LM}^{opt}\right] \\ &= [\alpha + (1-\alpha)\Psi]\Lambda\frac{\tau_c}{2}P(\hat{G}^*)^2\zeta_{h_D} \\ &\times \left[\left(1 - \frac{\pi^2}{16}\right)\sum_{n\in\mathcal{N}}\delta_n^2 + \frac{\pi^2}{16}\left(\sum_{n\in\mathcal{N}}\delta_n\right)^2\right].\end{aligned}$$

- Similarly,

$$\overline{P_D^{opt}} = P(\hat{G}^*)^2\zeta_{h_D}\left[\left(1 - \frac{\pi^2}{16}\right)\sum_{n\in\mathcal{N}}\delta_n^2 + \frac{\pi^2}{16}\left(\sum_{n\in\mathcal{N}}\delta_n\right)^2\right]. \quad (1)$$

Simulation: Harvested Energy

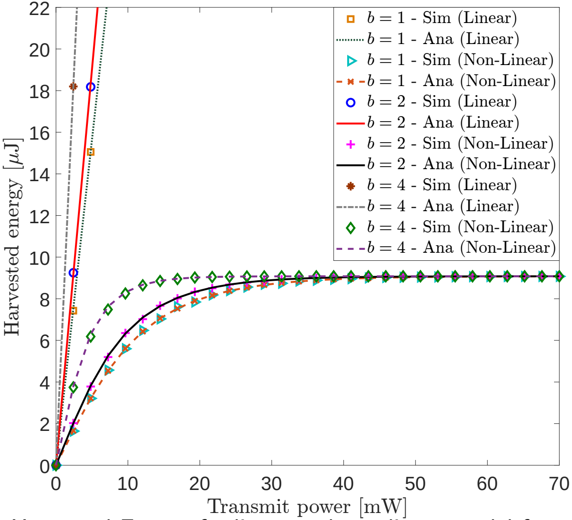


Figure: The Harvested Energy for linear and non linear model for $q \in \{1, 2, 4\}$.

Simulation: Average achievable rate

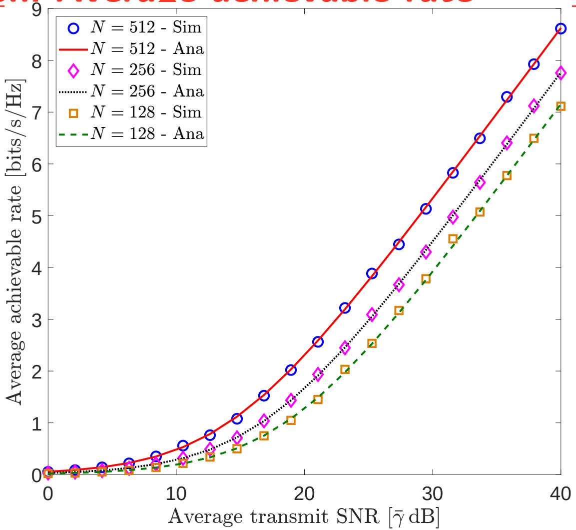
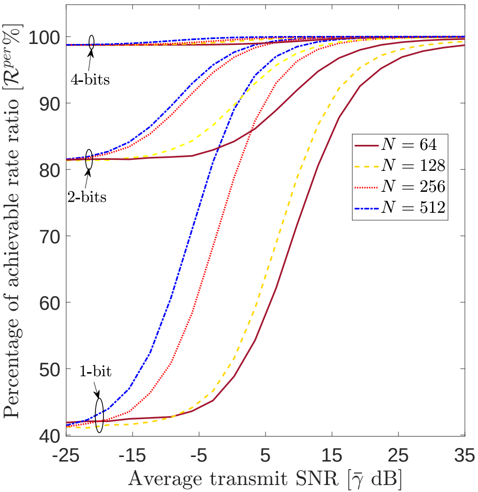


Figure: The average achievable rate for $N \in \{64, 128, 256, 512\}$.

Simulation: Phase-shift quantization



Conclusions

- The performance of an RIS-assisted relay system has been investigated.
- The optimal SNR that is attained through intelligent phase-shift controlling of the RIS elements has been probabilistically characterized by deriving a tight CDF approximation.
- Thereby, tight approximations/bounds for the fundamental performance metrics, including the average achievable rate, SNR/rate outage probability, and average SER have been derived.
- The optimal phase-shift matrix, which maximizes the end-to-end SNR has been derived, and thereby, maximizing the achievable rate-energy trade-off.
- The achievable rate, harvested energy and fundamental energy-rate trade-offs for the proposed RIS assisted relay systems for SWIPT has been derived.
- The accuracy of our analysis has been validated through the Monte-Carlo simulation.
- A rigorous set of numerical results has been presented to investigate the performance of the proposed RIS-assisted relay system.
- From our numerical results, we reveal that the RIS-assisted relay systems can enhance the end-to-end wireless communication performance.

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Thank you for your attention!

Questions ??

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