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Non-Linear Energy Harvesting in RIS-assisted URLLC Networks for Industry Automation

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Abstract—A reconfigurable intelligent surface (RIS)-assisted wireless communication system with non-linear energy harvesting (EH) and ultra-reliable low-latency constraints is considered for its possible applications in industrial automation. A distant data-center (DC) communicates with the multiple destination machines with the help of a full-duplex (FD) server machine (SM) and RIS. Assuming the deficiency of enough transmission power at the FD-SM, the SM is considered in the near vicinity of the destinations in the industry to forward the data received from the distant DC. The reception at SM is assisted by the RIS and a non-linear hybrid power-time splitting (PTS) based EH receiver architecture is adopted to extend the lifespan of SM, thus increasing network lifetime. The scheduling of multiple destinations is done by SM based on the considered selection criteria namely, random (RND) scheduling, absolute (ABS) channel-power-based (CPB) scheduling and normalized (NRM) CPB scheduling. The end-to-end performance of the considered FD RIS-assisted network is analyzed, and the expressions for the block error rate (BLER) for all scheduling schemes are derived. Moreover, the effects of number of RIS elements, packet size, channel uses on the system performance are analyzed for the considered ultra-reliable and low-latency communication (URLLC) network. The scheduling fairness of all the scheduling schemes is also analyzed to study the performance-fairness trade-off. The derived analytical results are verified through Monte-Carlo simulations.

Index Terms—Block error rate, non-linear energy harvesting, full-duplex, industry automation, reconfigurable intelligent surfaceS, scheduling, ultra-reliable low-latency communication.

I. INTRODUCTION

RECENT evolution in the wireless technologies has led to a significant demand for increased data-rates, connectivity and spectral efficiency. It is forecast that the number of connected devices would reach to 36048 million by 2025,

out of which 25 billion would be contributed by connections related to Internet-of-Things (IoT) which include connected cars, sensors, consumer-electronics, machines, etc. [1], [2]. It is expected that the peak data-rate and the user experience data-rate would reach to 1-20 Gbps and 10-100 Mbps, respectively [3]. Moreover, the spectral efficiency is expected to be $\times 1 - \times 3$, and the connection density is expected to grow to 10 k-10 m devices/km². Several technologies including ultra-dense networks, artificial intelligence (AI) empowered wireless networks, millimeter wave (mmWave), multiple-input multiple-output (MIMO) play a significant role in addressing the needs of the upcoming developments of the wireless technology [4]. For example, MIMO allows us to achieve a better spectral efficiency with the help of parallel multiple transmissions over the same bandwidth [5]. Large number of antenna deployment at the transmitting as well as receiving ends for high channel capacity is a major limitation of MIMO. This limitation can be addressed by introducing the energy efficient passive reconfigurable intelligent surfaces (RISs) which can be placed on various objects such as cars, buildings, etc. [4], [6]. The RIS technology has evolved recently as a promising technology with high potential to address the upcoming technology demands. RISs are the reconfigurable arrays of low-cost sub-wavelength sized passive elements (or meta-atoms) which can reflect the electromagnetic signals [7] and thus, they gain huge advantages over conventional relaying protocols. The electromagnetic properties of RISs can be engineered in order to achieve the adjustable magnitude and phase of the reflected signal. An RIS-assisted communication can be a cost and resource efficient technology that can support the communication in an IoT or smart industry (such as industry 4.0) environments. Considering its benefits, RIS is considered as an evolving technology for fifth generation (5G) and beyond 5G (B5G).

One of the main objectives of fifth 5G and B5G wireless communication systems is the energy efficiency (EE) (i.e., to reduce the power consumption without compromising the reliability performance of the system). In this aspect, energy harvesting (EH) techniques have evolved in the last few years as a green communication technology. The EH allows simultaneous wireless information and power transfer (SWIPT) in which both the energy as well as information are carried over the RF signals [8]–[10]. The EH can be classified as time-splitting (TS) EH and power-splitting (PS) EH. In the former, the EH is performed for a fraction of symbol duration, whereas the symbol is transmitted in the remaining fraction of duration using the harvested energy. Whereas, in the latter,

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a fraction of energy is harvested and remaining is used for information decoding (ID). In linear EH, the transmit power of the device linearly depends on the input power of the energy harvester, where the practical limitations of electrical circuitry, such as sensitivity and saturation, are not considered. Therefore, in [10], a practical non-linear TS EH full-duplex (FD) relay model was proposed and the outage performance as well as the quality-of-service (QoS) were analyzed. In non-linear EH, no energy is harvested till the input power at the harvester is less than a sensitivity value after which the harvester linearly harvests the energy till the input power reaches a saturation value. Similarly, [9] has presented a hybrid power-time splitting (PTS) EH at an FD relay and analyzed the block error rate (BLER), EE and system goodput for short-packet communication. In [8], SWIPT in relaying has been introduced for networks which operate with fixed block length messages. Along with efficient power utilization, effective spectrum utilization is also necessary, which can be achieved by using FD relaying. This is due to the fact that, unlike half-duplex (HD) where either reception or transmission occurs at a time, FD allows reception as well as transmission at the same time over the same frequency [11]. As the RISs can be reconfigured to achieve the constructive combining of reflected signals at the receiver, it may potentially reduce the effects of channel fading and self-interference (SI) at FD receivers through boosting the desired signal.

Another important factor that contributes to the practicality and improved performance of the wireless communication system is the multiple-user scenario, which has been considered in several recent works [11]–[13]. For example, in [12] absolute (ABS) and normalized (NRM) signal-to-noise (SNR) based scheduling schemes were studied with adaptive and affordable-rate transmission, and the expressions for average access time, average waiting time and average access rate were derived. Multiple user scheduling for FD two way relaying has been presented in [11], where the user pair is selected based on the max-min scheduling based on channel state information (CSI) and system state information (SSI). In [13] user selection based on channel norm and distance has been proposed for two way relaying communication along with physical network coding.

With focus on Industry 4.0 in recent years, the industrial automation has evolved as an important area of interest. The industrial automation leads to a requirement of controlling the machineries and devices by a distant controller, which makes the analysis of efficient communication involving EH, RIS and other methodologies for this purpose, becomes necessary. These applications demand high reliability and low-latency of the system, which can be improved by considering short-packet communications [8], [9], [14]. For example, [8] and [9] have used EH-enabled relaying for short-packet communication.

A. Related work

RIS-based communication has recently been explored by the research communities. For example, a point-to-point link for a base-station to user communication with RIS has been proposed in [15], where the system rate and energy effi-

ciency are optimized by varying the number of RIS meta-atoms. Recently, an indoor and outdoor environment based RIS assisted communication was proposed in [16], where the system model with a single RIS has been explored for both the scenarios. A two-hop hybrid RIS and relay based communication system was presented in [17], where the RIS assists the communication between source and relay in the first hop and relay to destination in the second hop. The authors have derived tight upper-bounds for achievable rates and analyzed the optimum transmission mode. A deep learning based indoor focussing based algorithm was proposed in [18]. An RIS assisted cognitive radio communication was studied in [19], where a single secondary user co-exists with multiple primary users. Several works have considered the communication networks for IoT and industry 4.0 [8], [20]–[22]. For example, in [20], polarization diversity for massive MIMO system was studied for industrial environment. An overview of the holographic MIMO surfaces (HMIMOS) was presented in [23], where the hardware architecture of reconfigurable surfaces, and the opportunities and important challenges for the HMIMOS design were discussed. The RIS-assisted communication has been analyzed for several scenarios like IoT, unmanned aerial vehicles (UAV), etc. [24]–[31]. An RIS-enabled IoT network was considered in [24], where using the Fisher-Snedecor F model, the shadowing and composite channels between a receiver-transmitter pair were analyzed. Another work that deploys RIS for IoT environment is [25], where sum mean squared error (SMSE) was minimized. The authors in [26] have introduced ambient backscatter communication assisted by RIS for short-range low-power IoT. To maximize the achievable rate in a UAV based system, an RIS-assisted communication for was proposed in [27].

RIS technology has been explored very recently for various wireless communication applications such as security [32]–[34], EH and wireless power transfer [35], [36], beamforming [4], non-orthogonal multiple access [37]–[39] and terminal positioning [40] etc. Recently, several works have also considered FD, SWIPT, ultra-reliable low-latency communications (URLLC) methods with RIS [8], [41]–[43]. The possible directions for the use of RIS technology in FD wireless communications have been discussed recently in [41]. They concluded that RISs can provide power-efficient support to FD transmission links for the devices which are out of the coverage of each other. However, the primary challenge of RIS-assisted FD communications is in the software design for the circuitry that controls the electromagnetic properties of the RIS. The URLLC constrained uplink communication for RIS-assisted system for sensor data transmission in industrial environment was presented in [42], and for mobile UAV system in [43].

Recently, several multi-user scenario involving RIS can be found in literature [44], [45]. For example, in [44] a multi-user scenario involving RIS was presented, where an energy efficiency (EE) maximization problem is addressed for optimization of RIS phase shifts and transmit power to maintain a certain quality of standard (QoS). Further, a deep-reinforcement learning based non-convex optimization algorithm for phase shifts is developed for the multiple-user

scenario in [45].

B. Motivations and contributions

Motivated by the benefits of RIS, EH, URLLC and multi-user diversity, and the need for analyzing the communication for industrial automation, in this paper, we consider an RIS-assisted wireless system with non-linear EH for industrial applications with URLLC. We consider a data-center (DC) which communicates with multiple destination machines with the help of an FD server machine (SM) in an industrial environment. We adopt a non-linear hybrid PTS EH model to harvest energy at the SM. *To the best of the authors' knowledge, none of the previous works [8], [41]–[43] have considered an RIS assisted FD-enabled hybrid PTS non-linear EH server-machine (SM) in URLLC with multiple destinations for industrial applications.* The major contributions of this paper are as follows.

- 1) RIS allows a controlled beamforming towards the desired receiver, improving the signal strength and quality of the received signal over longer distance, thus in this work, an RIS is used to assist the communication between the DC and SM in an industrial environment.
- 2) Industrial applications demand for efficient power utilization. To address this aspect, the SM adopts the non-linear hybrid PTS model for EH. For efficient spectrum utilization, the SM is considered to be operating in the FD mode of transmission.
- 3) Since the SM is responsible for scheduling of the destinations, we consider three scheduling schemes namely random scheduling, ABS channel-power-based (CPB) scheduling and NRM CPB scheduling schemes. Further, the end-to-end approximate closed-form expressions for the BLER and the resultant cumulative distribution function (CDF) for all the scheduling schemes are derived. Moreover, by considering the high-reliability aspects for industrial automation, short packet communication is used.
- 4) Finally, the performance of the system with non-linear EH is compared with the linear EH scenario for different number of RIS atoms. We further analyze the effect of various parameters, such as number of RIS meta-atoms, positioning of RIS, non-linear EH, packet size, blocklength, fairness of scheduling schemes, etc., on the system performance. To confirm the correctness of the analysis, the Monte-Carlo simulations are used.

Rest of the paper is organized as follows. The system model and non-linear hybrid EH scheme are described in Section II, while the performance analysis is presented in Section III. The numerical results are presented and discussed in Section IV. Finally, the conclusions and future scope are mentioned in Section V. The notations used in this paper are tabulated in Table I.

II. SYSTEM MODEL

In this paper, we consider an RIS-assisted wireless communication system in an industrial environment as shown in Fig. 1. A DC communicates with K number of destinations (D), but with one in the current time-slot, with the help of an RIS-assisted SM. The DC and D are assumed to operate in HD

mode, whereas SM operates in FD transmission mode. The DC comprises of one transmit antenna, whereas D has one receive antenna. The FD SM consists of two antennas, one each for reception from RIS and transmission to D . The FD transmission at the SM results in the residual self interference (RSI) between its transmitting and receiving antennas [46]. We consider the downlink communication to the destination machines, hence the DC comprises of single antenna [15]. The D is placed inside the industrial indoor environment, which receives the control signals from a distant controller placed in an outdoor environment. The RIS with large number of reflecting meta-atoms can achieve high beamforming gain as of massive MIMO with significantly low power consumption [4], assisting the communication by single-antenna possible. The SM acts as an intermediate communicating device that assists the reception of the information from the DC from outdoors and communicates it to the scheduled D placed in the industrial environment. The distance between the DC and SM, i.e. d_{sr} , is larger as compared to the distance between SM and k th D , i.e. d_{rk} , hence the signal transmitted by DC may attenuate significantly till it reaches the SM. Therefore, the reception at the SM is assisted by an RIS having N meta-atoms. The RIS assists reception at the SM, by forming a controlled beam towards SM and by controlling the reflection coefficient of each meta-atom. While the reflection coefficient of the n th meta-atom is assumed to be $\eta_n e^{j\theta_n}$, h_{sin} , h_{inr} and h_{rk} represent the channels from DC to n th RIS atom, n th RIS atom to SM and SM to k th D , respectively. For simplicity, it is assumed that the direct channels from DC to SM, and DC to D and RIS to D do not exist because of a poor channel conditions [47]. The DC to SM and the SM to destination transmissions take place simultaneously. The SM is responsible for scheduling of the destinations. Furthermore, the environment from DC to SM is outdoor and that from SM to destinations is indoor industrial environment. Considering these facts and larger distance (deep fading) between RIS to destinations, the signal strength received at destinations directly from RIS is significantly small and hence can be ignored. Furthermore, We assume the channels $|h_{pq}|$ to be Nakagami- m distributed with parameters m_{pq} and Ω_{pq} , whereas $(pq) \in \{(sin), (inr), (rk), (rr)\}$, and h_{rr} denotes the RSI channel. The Nakagami- m distribution is a generalized distribution and can be related to other distributions like Rayleigh, etc. by varying the value of m_{pq} . Recently, Nakagami- m distributed channels have been explored for RIS applications [48], [49]. Based on the scheduling schemes deployed, the resultant channel between SM and the scheduled D varies and is given as h_{rd} . Note that, for single user scenario, $(rk) = (rd)$. Therefore, the probability distribution function (PDF) and CDF respectively are given as

$$f_{|h_{pq}|}(x) = \frac{2m_{pq}^{m_{pq}}}{\Gamma(m_{pq})\Omega_{pq}^{m_{pq}}} x^{2m_{pq}-1} \exp\left(-\frac{m_{pq}}{\Omega_{pq}} x^2\right); x > 0, \quad (1)$$

$$\text{and, } F_{|h_{pq}|}(x) = \frac{\gamma\left(m_{pq}, \frac{m_{pq}}{\Omega_{pq}} x^2\right)}{\Gamma(m_{pq})}, \quad (2)$$

TABLE I: Definitions for Notations.

Sr. No.	Notation	Definition
1	DC	Data-center
2	D	Destination
3	SM	Server machine
4	h_{si_n}	Channel from DC to n th RIS atom
5	$h_{i_n r}$	Channel from n th RIS atom to SM
7	h_{sr}	Channel from DC to SM
8	h_{rk}	Channel from SM to k th D
9	P_s	Transmit power of DC
10	P_r	Transmit power of SM
11	m_{pq}	Shape parameter for the channel indicated by the pair pq , where $(pq) \in \{(si_n), (i_n r), (rk), (rr)\}$
12	Ω_{pq}	Channel power for channel indicated by the pair pq , where $(pq) \in \{(si_n), (i_n r), (rk), (rr)\}$
13	η_n	Reflection coefficient at each RIS atom
14	θ_n	Phase of reflection coefficient
15	N	Number of RIS atoms
16	w_r, w_d	AWGN noise for DC to SM, SM to D , respectively
17	$f(\cdot)$	Probability density function (PDF)
18	$F(\cdot)$	Cumulative distribution function (CDF)
19	M	Number of channels
20	μ_s	Number of channel uses
21	K	Number of destinations

Fig. 1: Considered RIS-assisted and non-linear EH based system model for its possible industry applications.

where m_{pq} is the shape parameter and Ω_{pq} is the channel power. Note that Ω_{pq} is determined using the distance dependent path-loss model. In order to efficiently use the limited amount of energy at the SM, the SM receiver considers a non-linear hybrid PTS EH.

The transmission takes place in a time-slotted fashion, such that each frame is divided into an initialization interval of L_i length symbols and M length symbols corresponding to transmission duration. The initialization interval is required for SWIPT parameter decision and blocklength allocation, whereas in the transmission duration, the DC transmits the signal, SM performs EH and the signal is received at D . The total duration of the frame is given as $(L_i + M)T_s$, where T_s is the symbol duration.

A. Received signal statistics

Let DC transmits the symbol $x_s[l]$ to the SM in the l th time slot with transmission power P_s . The signal received at the SM in the l th time slot is given as

$$y_r[l] = \sqrt{P_s} h_{sr} x_s[l] + \sqrt{P_r} h_{rr} x_r[l] + w_r, \quad (3)$$

where w_r is the additive white Gaussian noise (AWGN) with $w_r \sim \mathcal{N}(0, \sigma_r^2)$, $\sqrt{P_r} h_{rr} x_r[l]$ is the RSI such that the SM transmits $x_r[l]$ in l th time-slot with power P_r . Here, h_{sr} is the equivalent channel from DC to SM, considering RIS assisted reception at the SM and is given by

$$h_{sr} = \sum_{n=1}^N h_{si_n} \eta_n e^{j\theta_n} h_{i_n r}. \quad (4)$$

The channel coefficients can be represented as $h_{sin} = |h_{sin}|e^{j\theta_{sin}}$ and $h_{inr} = |h_{inr}|e^{j\theta_{inr}}$. As discussed previously, the beamforming is achieved by intelligently controlling the reflection coefficient. Therefore, to have optimal signal reception at SM and maximize the signal-to-interference-plus-noise ratio (SINR) of the DC to SM link, the resultant phase of h_{sr} must be zero, i.e., $\theta_n + \theta_{sin} + \theta_{inr} = 0$, or $\theta_n = -(\theta_{sin} + \theta_{inr})$ [4], [50]. Therefore, the channel coefficient h_{sr} for optimal SINR becomes

$$h_{sr} = \sum_{n=1}^N \eta_n |h_{sin}| |h_{inr}|. \quad (5)$$

From (5), it can be observed that h_{sr} is summation of large number of random variables, therefore, according to central limit theorem (CLT), $|h_{sr}|$ follows Gaussian distribution such that $|h_{sr}| \sim \mathcal{N}(\mu_h, \sigma_h^2)$, where the values of μ_h and σ_h^2 are given by

$$\mu_h = \eta_n \sum_{n=1}^N \left(\frac{\Omega_{sin} \Omega_{inr}}{m_{sin} m_{inr}} \right)^{\frac{1}{2}} \frac{\Gamma(m_{sin} + \frac{1}{2}) \Gamma(m_{inr} + \frac{1}{2})}{\Gamma(m_{sin}) \Gamma(m_{inr})}, \quad (6)$$

$$\sigma_h^2 = \eta_n^2 \sum_{n=1}^M \left[\Omega_{sin} \Omega_{inr} - \left(\left(\frac{\Omega_{sin} \Omega_{inr}}{m_{sin} m_{inr}} \right)^{\frac{1}{2}} \frac{\Gamma(m_{sin} + \frac{1}{2}) \Gamma(m_{inr} + \frac{1}{2})}{\Gamma(m_{sin}) \Gamma(m_{inr})} \right)^2 \right]. \quad (7)$$

The PDF and CDF of $|h_{sr}|$ are respectively given as

$$f_{|h_{sr}|}(x) = \frac{\alpha}{\sqrt{2\pi\sigma_h^2}} \exp\left(-\frac{(x - \mu_h)^2}{2\sigma_h^2}\right); \quad x > 0, \quad (8)$$

$$F_{|h_{sr}|}(x) = 1 - \alpha Q\left(\frac{x - \mu_h}{\sigma_h}\right), \quad (9)$$

where α is the normalization factor defined as $\alpha = \left[Q\left(\frac{\mu_h}{\sigma_h}\right) \right]^{-1}$ and $Q(\cdot)$ is the standard Q -function defined as $Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^\infty e^{-t^2/2} dt$. The signal received at scheduled D in l th time-slot is given as

$$y_d[l] = \sqrt{P_r} h_{rd} x_r[l] + w_d, \quad (10)$$

where $w_d \sim \mathcal{N}(0, \sigma_d^2)$ is the AWGN with variance σ_d^2 .

B. CSI estimation and central control unit

Recently, several works can be found that address the CSI estimation for RIS based communication systems [51]–[54]. We assume that the DC, SM and RIS are fully coordinated with the help of a central control unit (CCU). The CCU assists in gathering the CSI, such that the CSI is completely known at the RIS for tuning the phases of reflection coefficients. The CSI estimation is performed with techniques such as PARAllel FACtor (PARAFAC) decomposition based algorithms [51].

Fig. 2: P_r vs P_s for PS, TS and hybrid PTS EH.

C. Non-linear hybrid power-time splitting EH

The SM adopts a non-linear hybrid PTS EH. The traditional linear EH techniques lack the practicality of the energy harvester, that would not perform EH till the input power is less than a sensitivity value P_{min} as well as saturation P_{max} [9], [10]. In order to achieve the efficient EH, we adopt a hybrid PTS EH [9]. In PTS EH, the transmission phase M is divided into three phases, v , n_1 and n_2 . In the v , the SM performs EH using the TS protocol, in interval n_1 , the SM performs EH using the PS protocol with PS ratio ρ and ID with $(1 - \rho)$ fraction, where $\rho \in [0, 1]$, and for remaining n_2 interval, the SM performs the transmission from SM to desired D . The energy harvested by TS and PTS protocol is given as $\eta v T_s P_{in}$ and $\eta \rho n_1 T_s P_{in}$ respectively, where η is the power efficiency factor and the input power is given as $P_{in} = |h_{sr}|^2 P_s$. Therefore, the net energy harvested is given as $E_{PTS} = \eta (v T_s + \rho n_1 T_s) P_{in}$. The transmit power of SM is given as

$$P_r = \begin{cases} 0, & P_{in} < P_{min}, \\ \beta P_s |h_{sr}|^2, & P_{min} \leq P_{in} \leq P_{max}, \\ \beta P_{max}, & P_{in} > P_{max}, \end{cases} \quad (11)$$

where, $\beta = \frac{\eta(v T_s + \rho n_1 T_s)}{n_2 T_s}$. In order to analyze the efficiency of the EH protocols, we plot the P_r vs the P_s for TS, PS, and hybrid PTS EH is as shown in Fig. 2. It can be observed that highest EH is observed in case of Hybrid PTS EH as compared to TS EH and PS EH.

III. PERFORMANCE ANALYSIS

Using (3) and (10), the SINR for DC to SM link and SNR for the SM to D link are given as Υ_{sr} and Υ_{rd} , respectively, as

$$\Upsilon_{sr} = \frac{P_s |h_{sr}|^2}{P_r |h_{rr}|^2 + \sigma_r^2}, \quad (12)$$

$$\Upsilon_{rd} = \frac{P_r |h_{rd}|^2}{\sigma_d^2}. \quad (13)$$

Note that $|h_{rd}|$ is the equivalent channel between the SM and the selected D . Based on the values of P_r as in (11), Υ_{sr} and Υ_{rd} are written as [10]

$$\Upsilon_{sr} = \begin{cases} \frac{P_s |h_{sr}|^2}{\sigma_r^2}, & P_{in} < P_{min}, \\ \frac{1}{\beta |h_{rr}|^2}, & P_{min} \leq P_{in} \leq P_{max}, \\ \frac{P_s |h_{sr}|^2}{\beta P_{max} |h_{rr}|^2}, & P_{in} > P_{max}, \end{cases} \quad (14)$$

and

$$\Upsilon_{rd} = \begin{cases} 0, & P_{in} < P_{min}, \\ \frac{\beta P_s |h_{sr}|^2 |h_{rd}|^2}{\sigma_d^2}, & P_{min} \leq P_{in} \leq P_{max}, \\ \frac{\beta P_{max} |h_{rd}|^2}{\sigma_d^2}, & P_{in} > P_{max}, \end{cases} \quad (15)$$

respectively.

The $|h_{pq}|^2$ forms the gamma distribution with parameters m_{pq} and Ω_{pq} where $(pq) \in \{(si_n), (i_n r), (rk), (rr)\}$.

The PDF and CDF for $|h_{pq}|^2$ are respectively given as

$$f_{|h_{pq}|^2}(x) = \frac{m_{pq}^{m_{pq}}}{\Gamma(m_{pq})\Omega_{pq}^{m_{pq}}} x^{m_{pq}-1} \exp\left(-\frac{m_{pq}}{\Omega_{pq}}x\right), \quad (16)$$

and

$$F_{|h_{pq}|^2}(x) = \frac{\gamma\left(m_{pq}, \frac{m_{pq}}{\Omega_{pq}}x\right)}{\Gamma(m_{pq})}. \quad (17)$$

Similarly, the CDF and PDF of $|h_{sr}|^2$ are respectively given as [50]

$$F_{|h_{sr}|^2}(x) = 1 - \alpha Q\left(\frac{\sqrt{x} - \mu_h}{\sigma_h}\right), \quad (18)$$

$$f_{|h_{sr}|^2}(x) = \frac{\alpha}{2\sqrt{2\pi}\sigma_h^2 x} \exp\left(-\frac{(\sqrt{x} - \mu_h)^2}{2\sigma_h^2}\right). \quad (19)$$

A. CDF of end-to-end SINR

For the BLER analysis, first we find the CDF of the end-to-end SINR. The end-to-end SINR is mathematically calculated as

$$\Upsilon = \min(\Upsilon_{sr}, \Upsilon_{rd}). \quad (20)$$

The end-to-end SINR's CDF ($F_\Upsilon(\Upsilon_{th})$) is given as

$$\begin{aligned} F_\Upsilon(\Upsilon_{th}) &= P(\min(\Upsilon_{sr}, \Upsilon_{rd}) < \Upsilon_{th}), \\ &= 1 - P(\Upsilon_{sr} \geq \Upsilon_{th} \cap \Upsilon_{rd} \geq \Upsilon_{th}), \\ &= 1 - P((P_{in} \geq P_{min}) \cap (\Upsilon_{sr} > \Upsilon_{th}) \cap (\Upsilon_{rd} > \Upsilon_{th})), \\ &= 1 - I_1 - I_2, \end{aligned} \quad (21)$$

where I_1 and I_2 are respectively defined as

$$\begin{aligned} I_1 &= P(P_{min} \leq P_{in} \leq P_{max} \cap \Upsilon_{sr} > \Upsilon_{th} \cap \Upsilon_{rd} > \Upsilon_{th}), \\ &= \begin{cases} 0, & |h_{rd}|^2 < \frac{D}{C}, \\ P(|h_{rr}|^2 < A) P\left(\frac{D}{|h_{rd}|^2} < |h_{sr}|^2 \leq C\right), & \frac{D}{C} < |h_{rd}|^2 < \frac{D}{B}, \\ P(|h_{rr}|^2 < A) P\left(B < |h_{sr}|^2 \leq C\right), & |h_{rd}|^2 > \frac{D}{B}, \end{cases} \end{aligned} \quad (22)$$

and

$$\begin{aligned} I_2 &= P(P_{in} \geq P_{max} \cap \Upsilon_{sr} > \Upsilon_{th} \cap \Upsilon_{rd} > \Upsilon_{th}), \\ &= \begin{cases} P(|h_{rd}|^2 > E) P(|h_{sr}|^2 > C), & |h_{rr}|^2 < A, \\ P(|h_{rd}|^2 > E) P\left(|h_{sr}|^2 > \frac{C}{A} |h_{rr}|^2\right), & |h_{rr}|^2 > A, \end{cases} \end{aligned} \quad (23)$$

where $A = \frac{1}{\beta \Upsilon_{th}}$, $B = \frac{P_{min}}{P_s}$, $C = \frac{P_{max}}{P_s}$, $D = \frac{\sigma_d^2 \Upsilon_{th}}{\beta P_s}$ and $E = \frac{\sigma_d^2 \Upsilon_{th}}{\beta P_{max}}$. Further, we derive the approximate closed-form expressions for I_1 and I_2 as

$$\begin{aligned} I_1 &= \int_0^A f_{|h_{rr}|^2}(x) dx \int_{\frac{D}{C}}^{\frac{D}{B}} \left(f_{|h_{rd}|^2}(x) \int_{\frac{D}{x}}^C f_{|h_{sr}|^2}(y) dy \right) dx \\ &\quad + \int_0^A f_{|h_{rr}|^2}(x) dx \int_B^C f_{|h_{sr}|^2}(x) dx \int_{\frac{D}{B}}^\infty f_{|h_{rd}|^2}(x) dx, \end{aligned} \quad (24)$$

$$\begin{aligned} I_1 &= F_{|h_{rr}|^2}(A) \times \left[I_3 + \left(F_{|h_{sr}|^2}(C) - F_{|h_{sr}|^2}(B) \right) \right. \\ &\quad \left. \times \left(1 - F_{|h_{rd}|^2}\left(\frac{D}{B}\right) \right) \right], \end{aligned} \quad (25)$$

where

$$\begin{aligned} I_3 &= \int_{\frac{D}{C}}^{\frac{D}{B}} \left(f_{|h_{rd}|^2}(x) \int_{\frac{D}{x}}^C f_{|h_{sr}|^2}(y) dy \right) dx, \\ &= \int_{\frac{D}{C}}^{\frac{D}{B}} f_{|h_{rd}|^2}(x) \left[F_{|h_{sr}|^2}(C) - F_{|h_{sr}|^2}\left(\frac{D}{x}\right) \right] dx, \\ &= F_{|h_{sr}|^2}(C) \left(F_{|h_{rd}|^2}\left(\frac{D}{B}\right) - F_{|h_{rd}|^2}\left(\frac{D}{C}\right) \right) - I_4. \end{aligned} \quad (26)$$

The term I_4 in (26) is defined as

$$I_4 = \int_{\frac{D}{C}}^{\frac{D}{B}} f_{|h_{rd}|^2}(x) F_{|h_{sr}|^2}\left(\frac{D}{x}\right) dx. \quad (27)$$

We use the Gaussian-Chebyshev-Quadrature (GCQ) method to approximate (27) [55, eq. 25.4.30] as

$$I_4 = \frac{\frac{D}{B} + \frac{D}{C}}{2} \sum_{g=1}^G \frac{\pi}{G} \sqrt{1 - \phi_g^2} f_{|h_{rd}|^2}(z_1) F_{|h_{sr}|^2}\left(\frac{D}{z_1}\right) + R_G, \quad (28)$$

where $\phi_g = \cos\left(\frac{2g-1}{2G}\pi\right)$, $z_1 = \left(\frac{\frac{D}{B}-\frac{D}{C}}{2}\right)\phi_g + \left(\frac{\frac{D}{B}+\frac{D}{C}}{2}\right)$, G is the complexity-accuracy trade-off factor and R_G is the error term that can be neglected at sufficiently higher value of G . The final approximate closed-form expression for I_1 is given in (29), shown on the top of the next page.

Similarly, the expression for I_2 is given as

$$\begin{aligned} I_2 &= \int_E^\infty f_{|h_{sr}|^2}(x) dx \int_A^\infty \left(f_{|h_{rr}|^2}(x) \int_{\frac{C}{A}x}^\infty f_{|h_{sr}|^2}(y) dy \right) dx \\ &\quad + \int_E^\infty f_{|h_{sr}|^2}(x) dx \int_C^\infty f_{|h_{sr}|^2}(x) dx \int_0^A f_{|h_{rr}|^2}(x) dx, \end{aligned} \quad (30)$$

$$I_2 = \left(1 - F_{|h_{rd}|^2}(E) \right) \times \left[I_5 + \left(1 - F_{|h_{sr}|^2}(C) \right) F_{|h_{rr}|^2}(A) \right], \quad (31)$$

where

$$\begin{aligned} I_5 &= \int_A^\infty \left(f_{|h_{rr}|^2}(x) \int_{\frac{C}{A}x}^\infty f_{|h_{sr}|^2}(y) dy \right) dx, \\ &= 1 - \int_0^A f_{|h_{rr}|^2}(x) \left(1 - \int_0^{\frac{C}{A}x} f_{|h_{sr}|^2}(y) dy \right) dx, \\ &= 1 - F_{|h_{rr}|^2}(A) + I_6. \end{aligned} \quad (32)$$

$$I_1 = F_{|h_{rr}|^2}(A) \times \left[F_{|h_{sr}|^2}(C) \left(F_{|h_{rd}|^2}\left(\frac{D}{B}\right) - F_{|h_{rd}|^2}\left(\frac{D}{C}\right) \right) - \left(\frac{\frac{D}{B} + \frac{D}{C}}{2} \sum_{g=1}^G \frac{\pi}{G} \sqrt{1 - \phi_g^2} f_{|h_{rd}|^2}(z_1) F_{|h_{sr}|^2}\left(\frac{D}{z_1}\right) + R_G \right) \right. \\ \left. + \left(F_{|h_{sr}|^2}(C) - F_{|h_{sr}|^2}(B) \right) \times \left(1 - F_{|h_{rd}|^2}\left(\frac{D}{B}\right) \right) \right]. \quad (29)$$

Here I_6 in (32) is defined as

$$I_6 = \int_0^A f_{|h_{rr}|^2}(x) \left(\int_0^{\frac{C}{A}x} f_{|h_{sr}|^2}(y) dy \right) dx, \\ = \int_0^A f_{|h_{rr}|^2}(x) F_{|h_{sr}|^2}\left(\frac{C}{A}x\right) dx. \quad (33)$$

Substituting $x = Ay$ in (33), we get

$$I_6 = A \int_0^1 f_{|h_{rr}|^2}(Ay) F_{|h_{sr}|^2}(Cy) dy. \quad (34)$$

I_6 in (34) is approximated using the GCQ method as [55]

$$I_6 = A \sum_{l_1=1}^L \frac{\pi}{2L} \sqrt{1 - \phi_{l_1}^2} f_{|h_{rr}|^2}(Az_2) F_{|h_{sr}|^2}(Cz_2) + R_L, \quad (35)$$

where $\phi_{l_1} = \cos\left(\frac{2l_1-1}{2L}\pi\right)$, $z_2 = \left(\frac{\phi_{l_1}+1}{2}\right)$, L is the complexity-accuracy trade-off factor and R_L is the error term that can be neglected at sufficiently higher values of L . The final approximate closed-form expression for I_2 is given in (36), shown on the bottom of this page. Finally, the CDF can be obtained by substituting (29) and (36) in (21).

B. Multi-destination Scheduling

In order to transmit the data to the destination, the scheduling is performed at the SM. Three scheduling schemes, namely RND scheduling, ABS CPB scheduling and NRM CPB scheduling are considered in this work. As the destination selection is done at the SM the channel from SM to D would be affected and based on the scheduling scheme, $F_{|h_{rd}|^2}$ and $f_{|h_{rd}|^2}$ in (29) and (36) would change.

1) *Random (RND) Scheduling*: In RND scheduling, the SM randomly selects a destination for the data transmission. The probability that a destination is selected randomly is $\frac{1}{K}$. The CDF and PDF of $|h_{rd}|^2$ are respectively given as

$$F_{|h_{rd}|^2}(x) = \frac{1}{K} \sum_{k=1}^K F_{|h_{rk}|^2}(x), \quad (37)$$

$$f_{|h_{rd}|^2}(x) = \frac{1}{K} \sum_{k=1}^K f_{|h_{rk}|^2}(x). \quad (38)$$

The access probability of a user for RND scheduling scheme is given as

$$P_k = \frac{1}{K}. \quad (39)$$

2) *Absolute (ABS) channel-power-based (CPB) Scheduling*: The destination with maximum channel power is selected in ABS CPB scheduling scheme. Therefore, the CDF for $|h_{rd}|^2$, when k th destination is selected, is given as

$$F_{|h_{rd}|^2}(x) = \sum_{k=1}^K \Pr\left(|h_{rk}|^2 < x \cap |h_{rj}|^2 < |h_{rk}|^2\right), \\ \forall j \in \{1, 2, \dots, K\} \text{ and } j \neq k, \\ = \sum_{k=1}^K \int_0^x f_{|h_{rk}|^2}(y) \prod_{\substack{j=1 \\ j \neq k}}^K F_{|h_{rj}|^2}(y) dy. \quad (40)$$

Similarly, the PDF is given as

$$f_{|h_{rd}|^2}(x) = \sum_{k=1}^K \left(f_{|h_{rk}|^2}(x) \prod_{\substack{j=1 \\ j \neq k}}^K F_{|h_{rj}|^2}(x) \right). \quad (41)$$

The probability that the k th destination is accessed for ABS CPB scheduling scheme is given as

$$P_k = \Pr\left(|h_{rk}|^2 > |h_{rj}|^2\right), \quad \forall j \in \{1, 2, \dots, K\} \text{ and } j \neq k, \\ = \int_0^\infty f_{|h_{rk}|^2}(x) \prod_{\substack{j=1 \\ j \neq k}}^K \int_0^x f_{|h_{rj}|^2}(y) dy dx, \\ = \int_0^\infty f_{|h_{rk}|^2}(x) \prod_{\substack{j=1 \\ j \neq k}}^K F_{|h_{rj}|^2}(x) dx. \quad (42)$$

Substituting the PDF and CDF of $|h_{rk}|^2$ and $|h_{rj}|^2$ as in (16) and (17) respectively and substituting $\exp\left(-\frac{m_{rk}}{\Omega_{rk}}x\right) = t$ in (42), we have

$$P_k = \int_0^1 b_k (\ln(t))^{m_{rk}-1} \prod_{\substack{j=1 \\ j \neq k}}^K \frac{\gamma(m_{rj}, -a_j \ln(t))}{\Gamma(m_{rj})}, \quad (43)$$

$$I_2 = \left(1 - F_{|h_{rd}|^2}(E) \right) \times \left[1 - F_{|h_{rr}|^2}(A) + \left(A \sum_{l=1}^L \frac{\pi}{2L} \sqrt{1 - \phi_l^2} f_{|h_{rr}|^2}(Az_2) F_{|h_{sr}|^2}(Cz_2) + R_L \right) \right. \\ \left. + \left(1 - F_{|h_{sr}|^2}(C) \right) F_{|h_{rr}|^2}(A) \right]. \quad (36)$$

where, $b_k = \frac{(-1)^{m_{rk}-1}}{\Gamma(m_{rk})}$ and $a_j = \frac{m_{rj}}{m_{rk}} \frac{\Omega_{rk}}{\Omega_{rj}}$. The approximate closed-form expression can be obtained using the GCQ method [55, eq. 25.4.30]

$$P_k = b_k \frac{\pi}{2\hat{G}} \sum_{\hat{g}=1}^{\hat{G}} \sqrt{1 - \phi_{\hat{g}}^2} (\ln(z_g))^{m_{rk}-1} \times \prod_{\substack{j=1 \\ j \neq k}}^K \frac{\gamma(m_{rj}, -a_j \ln(z_g))}{\Gamma(m_{rj})} + R_{\hat{G}}, \quad (44)$$

where $\phi_{\hat{g}} = \cos\left(\frac{2\hat{g}-1}{2\hat{G}}\pi\right)$, $z_g = \left(\frac{\phi_{\hat{g}}+1}{2}\right)$, $\hat{G}$ is the complexity-accuracy trade-off factor and $R_{\hat{G}}$ is the error term that can be neglected at sufficiently higher value of $\hat{G}$.

3) *Normalized (NRM) channel-power-based (CPB) Scheduling*: Since ABS CPB scheduling scheme always selects the destination with maximum channel power, it is an unfair scheduling scheme. To address this, we use the NRM CPB scheduling scheme, where the destination with maximum normalized channel power is selected [12]. The CDF of $|h_{rd}|^2$, when k th destination is selected, is given as

$$\begin{aligned} F_{|h_{rd}|^2}(x) &= \sum_{k=1}^K \Pr\left(|h_{rk}|^2 < x \cap \frac{|h_{rj}|^2}{\Omega_{rj}} < \frac{|h_{rk}|^2}{\Omega_{rk}}\right), \\ &= \sum_{k=1}^K \Pr\left(|h_{rk}|^2 < x \cap |h_{rj}|^2 < |h_{rk}|^2 \frac{\Omega_{rj}}{\Omega_{rk}}\right), \\ &\quad \forall j \in \{1, 2, \dots, K\} \text{ and } j \neq k, \\ &= \sum_{k=1}^K \int_0^x f_{|h_{rk}|^2}(y) \prod_{\substack{j=1 \\ j \neq k}}^K F_{|h_{rj}|^2}\left(\frac{\Omega_{rj}}{\Omega_{rk}}y\right) dy. \end{aligned} \quad (45)$$

Similarly, the PDF is given as

$$f_{|h_{rd}|^2}(x) = \sum_{k=1}^K \left(f_{|h_{rk}|^2}(x) \prod_{\substack{j=1 \\ j \neq k}}^K F_{|h_{rj}|^2}\left(\frac{\Omega_{rj}}{\Omega_{rk}}x\right) \right). \quad (46)$$

The access probability of the k th destination for the NRM CPB scheduling scheme is given as

$$\begin{aligned} P_k &= \Pr\left(\frac{|h_{rk}|^2}{\Omega_{rk}} > \frac{|h_{rj}|^2}{\Omega_{rj}}\right), \\ &= \Pr\left(|h_{rk}|^2 > |h_{rj}|^2 \frac{\Omega_{rk}}{\Omega_{rj}}\right), \forall j \in \{1, 2, \dots, K\} \text{ and } j \neq k, \\ &= \int_0^\infty f_{|h_{rk}|^2}(x) \prod_{\substack{j=1 \\ j \neq k}}^K \int_0^{x \frac{\Omega_{rk}}{\Omega_{rj}}} f_{|h_{rj}|^2}(y) dy dx, \\ &= \int_0^\infty f_{|h_{rk}|^2}(x) \prod_{\substack{j=1 \\ j \neq k}}^K F_{|h_{rj}|^2}\left(\frac{\Omega_{rk}}{\Omega_{rj}}x\right) dx. \end{aligned} \quad (47)$$

By substituting the PDF and CDF of $|h_{rk}|^2$ and $|h_{rj}|^2$ as in (16) and (17) respectively and substituting $\exp\left(-\frac{m_{rk}}{\Omega_{rk}}x\right) = t$ in (47), we get

$$P_k = \int_0^1 b_k (\ln(t))^{m_{rk}-1} \prod_{\substack{j=1 \\ j \neq k}}^K \frac{\gamma(m_{rj}, -c_j \ln(t))}{\Gamma(m_{rj})}, \quad (48)$$

where $c_j = \frac{m_{rj}}{m_{rk}}$. The approximate closed-form expression can be obtained by using the GCQ method [55, eq. 25.4.30]

$$P_k = b_k \frac{\pi}{2\hat{L}} \sum_{\hat{l}=1}^{\hat{L}} \sqrt{1 - \phi_{\hat{l}}^2} (\ln(z_l))^{m_{rk}-1} \times \prod_{\substack{j=1 \\ j \neq k}}^K \frac{\gamma(m_{rj}, -c_j \ln(z_l))}{\Gamma(m_{rj})} + R_{\hat{L}}, \quad (49)$$

where $\phi_{\hat{l}} = \cos\left(\frac{2\hat{l}-1}{2\hat{L}}\pi\right)$, $z_l = \left(\frac{\phi_{\hat{l}}+1}{2}\right)$, $\hat{L}$ is the complexity-accuracy trade-off factor and $R_{\hat{L}}$ is the error term that can be neglected at sufficiently higher value of $\hat{L}$.

C. Fairness Analysis

The fairness of a scheduling scheme indicates how unbiased the scheme is while selection of the destination. The self-fairness for the k th D and the average fairness of the system are respectively given as [12]

$$s_k = -\frac{\log(P_k)}{\log(K)}, \quad (50)$$

$$\bar{s}_k = -\sum_{k=1}^K P_k s_k = -\sum_{k=1}^K P_k \frac{\log(P_k)}{\log(K)}, \quad (51)$$

where P_k is the access probability which can be obtained as in (39), (44) and (49) for RND, ABS CPB and NRM CPB scheduling schemes, respectively.

D. Block Error Rate

Considering the ϵ_{sd} as the overall BLER of the considered an RIS-assisted wireless network. It is assumed that the DC transmits b_s number of bits over M channels. For μ_s number of channels uses, the coding rate (bits/channel uses) is given as $r_{sd} = \frac{b_s}{\mu_s}$. Therefore, the overall BLER is

$$\epsilon_{sd}(r_{sd}) \approx \mathbb{E}\left\{Q\left(\frac{C(Y) - r_{sd}}{\sqrt{V(Y)/\mu_s}}\right)\right\}, \quad (52)$$

where $C(Y) = \log_2(1+Y)$ is the Shannon capacity and $V(Y) = \left(1 - (1+Y)^{-2}\right)(\log_2 e)^2$ is the channel dispersion that measures the stochastic channel variability relative to a deterministic channel with same channel capacity.

Due to mathematical intractability of the $\epsilon_{sd}(r_{sd})$ to obtain a closed-form expression, we adopt the linear approximation of the $\Xi = Q\left(\frac{C(Y) - r_{sd}}{\sqrt{V(Y)/\mu_s}}\right)$ as [56]

$$\Xi = \begin{cases} 1, & Y \leq \zeta, \\ 0.5 - \vartheta \sqrt{\mu_s} (Y - \theta), & \zeta < Y < \xi, \\ 0, & Y \geq \xi, \end{cases} \quad (53)$$

TABLE II: Simulation Parameters

Notation	Definition	Value
d_{si}	Distance from DC to RIS	95 m
d_{ir}	Distance from RIS to SM	95 m
d_{sr}	Distance from DC to SM	150 m
d_{rd}	Distance from SM to D	20 m
T_s	Symbol duration	1 μ s
N	Number of RIS	64
M	Number of channels	1000
ν	Number of intervals for TS EH	50
n_1	Number of intervals for PS EH	650
n_2	Number of intervals for transmission from SM to D	300
η_n	Reflection coefficient	0.2
ρ	Power splitting factor	0.6
η	Power efficiency	0.7
P_{min}	Sensitivity	0.5 W
P_{max}	Saturation	9 W
Ω_{rr}	RSI Power	-10 dB
m_{rr}	Shape parameter for RSI channel	3
m_{sin} m_{inr} m_{rk}	Shape parameters (except RSI)	1

where $\vartheta = \frac{1}{2\pi\sqrt{2^{r_{sd}}-1}}$, $\theta = 2^{r_{sd}} - 1$, $\zeta = \theta - \frac{1}{2\vartheta\sqrt{\mu_s}}$ and $\xi = \theta + \frac{1}{2\vartheta\sqrt{\mu_s}}$. The final BLER is obtained by averaging Ξ with respect to the PDF of Y . Therefore, we get

$$\epsilon_{sd}(r_{sd}) \approx \int_0^\infty \Xi(x) f_Y(x) dx \approx [\Xi(x) F_Y(x)]_0^\infty - \int_0^\infty F_Y(x) d\Xi(x), \quad (54)$$

where $F_Y(x)$ and $f_Y(x)$ are the CDF and PDF of Y respectively. Therefore, we have

$$\epsilon_{sd}(r_{sd}) = \vartheta\sqrt{\mu_s} \int_\zeta^\xi F_Y(x) dx. \quad (55)$$

In order to obtain the final expression of ϵ_{sd} , we apply the GCQ method [55, eq. 25.4.30]. We have

$$\epsilon_{sd}(r_{sd}) = \vartheta\sqrt{\mu_s} \left(\frac{\xi + \zeta}{2} \right) \sum_{p=1}^P \frac{\pi}{P} \sqrt{1 - \phi_p^2} F_Y(z_3) + R_P, \quad (56)$$

where $\phi_p = \cos\left(\frac{2p-1}{2P}\pi\right)$, $z_3 = \left(\frac{\xi-\zeta}{2}\right)\phi_p + \left(\frac{\xi+\zeta}{2}\right)$, P is the complexity-accuracy trade-off factor and R_P is the error term, being neglected for significantly higher value of P .

IV. NUMERICAL RESULTS AND DISCUSSION

Monte-Carlo simulations (MC) are used for verification of the presented numerical analysis. Unless specified otherwise, the symbol duration T_s is set as 1μ s, $M = 1000$, $N = 64$, $n_1 = 650$, $n_2 = 300$, $\nu = 50$, $\sigma_r^2 = \sigma_d^2 = \sigma^2 = 0$ dB, $P_{min} = 0.5$ W, $P_{max} = 9$ W, $d_{sr} = 150$ m, $d_{si} = 95$ m, $d_{ir} = 95$ m, $d_{rd} = 20$ m, $\eta_n = 0.2$, and RSI power $\Omega_{rr} = -10$ dB. The shape parameters for the RSI channel, $m_{rr} = 3$ and $m_{sin} =$

Fig. 3: Positioning of RIS

(a) BLER vs $\frac{P_s}{\sigma^2}$ (b) BLER vs $\frac{P_s}{\sigma^2}$

Fig. 4: (a) BLER vs $\frac{P_s}{\sigma^2}$ for different number of RIS atoms compared with no RIS case. (b) BLER vs $\frac{P_s}{\sigma^2}$ for different number of RIS atoms and linear and non-linear EH.

$m_{inr} = m_{rd} = 1$. The positioning of RIS follows $d_{sr} < \hat{d}_{si} + \hat{d}_{ir}$ as shown in Fig. 3. We consider the path-loss model in dB as $\Omega = \Omega_o + 10\nu \log_{10}(d)$, where $\nu = 3.5$ is the path-loss exponent for outdoor environment (for channels between DC to SM) and $\nu = 2.7$ is the path-loss exponent for indoor environment (for channel from SM to D), Ω_o is the reference path-loss which is considered as 35 dB and 25 dB for the DC to SM and SM to D links, respectively and d is the distance. The summary of the simulation parameters is tabulated in Table II.

Fig. 5: BLER vs $\hat{d}_{si}$ for different value of $\hat{d}$.

A. Impact of the number of RIS atoms (N)

We begin with analyzing the impact of the number of RIS atoms (N) on the system's performance as shown in Fig. 4a. It can be observed that when the value of N increases, the performance of the system improves and it requires less $\frac{P_s}{\sigma^2}$. Moreover, the BLER for given N attains approximately a common saturation. For example: when $\frac{P_s}{\sigma^2} \geq 12$ dBm and $N = 128$, the BLER starts saturating. The saturation is obtained at lower values of $\frac{P_s}{\sigma^2}$, as N increases which indicates that same BLER value can be obtained at lower transmit SNR if the number of RIS atoms is sufficiently increased. We further compare the performance with the direct transmission case, where no RIS is placed between the DC and SM. It can be concluded that inclusion of RIS between DC and SM significantly improves the BLER from 0.06 to a value less than 10^{-4} , allowing transmission of control signals from a longer distance to the controlling SM. The Fig. 4b shows a comparison between the non-linear and linear EH for different RIS atom. It can be observed that the linear EH leads to better reliability as compared to that for the linear EH, but non-linear EH is more practical with respect to the electronic circuitry of the harvester. The BLER plots for both approximately overlap for a particular region, which corresponds to the linear region where the input power of the harvester is between sensitivity and saturation of the harvester. In order to achieve same performance for non-linear EH compared to linear EH, the system has to be operated in linear region of harvester.

B. Impact of the RIS position

Another factor that affects the system's performance is the positioning of RIS. In order to analyze its effect, the BLER performance with respect to the position of RIS ($\hat{d}_{si}$) from DC is observed. We set $d_{si} = \sqrt{\hat{d}_{si}^2 + \hat{d}^2}$, where $\hat{d}$ is the perpendicular displacement of the RIS from the shortest path as illustrated in Fig. 3. Fig. 5 shows that irrespective of the values of $\hat{d}$, the optimum performance of the system is observed if the RIS is positioned either near to DC or SM.

Fig. 6: Effect of RIS on energy harvesting.

Fig. 7: BLER as a function of P_{max} for different values of N .

The symmetry in the curve is a result of the fact that as the distance of RIS increases from DC, it increases from SM with same factor, leading to the effective path-loss of the DC to SM channel same for $\hat{d}_{si} = d_o$ and $\hat{d}_{ir} = d_o$, where d_o is some distance between 0 to d_{sr} .

C. Impact of the RIS on EH

Further, we analyze the impact of the RIS on EH. Fig. 6 shows the transmit power of SM (P_r) versus the transmit power of the DC (P_s). It can be observed that for higher values of N and η_n , the amount of harvested energy is higher as compared to the direct transmission between DC and SM and low N and η_n case. It can be concluded that inclusion of RIS improves the EH amount significantly. Furthermore, it can be observed that, the value of P_r saturates to a common value due to non-linear EH for high value of η_n and N .

D. Impact of the saturation (P_{max})

The variation of the BLER with respect to P_{max} is as shown in Fig. 7. It was observed that the BLER decreases with increase in P_{max} as more power is available for EH and hence for transmission.

Fig. 8: $\frac{P_s}{\sigma^2}$ vs BLER for different value of n_1 , n_2 and v .

Fig. 10: Effect of the efficiency η and PS factor ρ .

Fig. 9: Effect of the number of bits b_s and channel uses μ_s on the BLER.

E. Effect of EH on the BLER

The effect of EH on the BLER is examined in Fig. 8 by varying the values of n_1 , n_2 and v . Initially, we keep the TS duration constant at $v = 50$ and vary n_1 and n_2 . The performance improves for the case where $n_1 > n_2$ due to higher time availability for ID. Similarly, for constant value of transmit duration ($n_2 = 300$), we can observe that the improved performance is obtained for $n_1 > v$. Further, we analyze the case of PS EH and TS EH. PS EH is achieved when the TS duration $v = 0$, whereas TS EH is achieved when the PS factor $n_1 = 0$. It can be observed from Fig. 8 that the PS EH ($v = 0$, $n_1 = 500$, $n_2 = 500$) leads to slightly better performance as compared to TS EH ($v = 500$, $n_1 = 0$, $n_2 = 500$).

F. Trade-off between the number of bits b_s and channel uses μ_s

The trade-off between the bits/channel (b_s) and the channel uses (μ_s) is observed in Fig. 9. It can be observed that the BLER performance improves for low values of b_s and high values of μ_s . The region for obtaining the optimal BLER lies

Fig. 11: BLER for different scheduling schemes and number of users.

$b_s \in [50, 100]$ and $\mu_s \in [600, 900]$. In Fig. 10, the trade-off between the power efficiency factor (η) and PS ratio (ρ) is illustrated. We can observe that the optimal performance is obtained at high values of both η and ρ .

G. Performance Comparison of Scheduling Schemes

In order to compare different scheduling schemes, we plot the BLER vs $\frac{P_s}{\sigma^2}$ for different values of number of users K and scheduling schemes as shown in Fig. 11. As can be seen that the BLER performance for all the scheduling schemes improves when K increases. It can be observed that the RND scheduling scheme performs worst, followed by NRM CPB and ABS CPB. Further, the fairness of the scheduling schemes is examined in Fig. 12. It can be observed that the ABS CPB results in a lower average fairness as compared to the NRM CPB and the RND scheduling schemes.

V. CONCLUSIONS AND FUTURE WORKS

In this paper, we analyzed the reliability performance of an RIS-assisted and non-linear EH based wireless communication

Fig. 12: Average fairness versus the number of destinations.

systems for industrial applications with URLLC constraints. We derived the end-to-end approximate closed-form expressions for the BLER and the equivalent CDF for the RND, ABS-CPB and NRM-CPB scheduling schemes. Further, we analyzed the effect of various parameters, such as number of RIS meta-atoms, blocklength, packet size and position of RIS on the system performance. Furthermore, the effect of RIS atoms on the EH was also demonstrated and noticed that the hybrid PTS EH does not guarantee the optimal performance, but leads to maximum EH. Moreover, it was observed that the BLER performance is better when the PS duration is higher as compared to the time splitting and transmission interval. The trade-off between the bits per channel and channel uses was analyzed and the optimal BLER region was determined. Moreover, we observed that for higher values of efficiency factor and PS ratio, the system becomes more reliable. The fairness of the scheduling schemes was analyzed, and it was observed that the NRM CPB and RND scheduling schemes are comparatively more fair as compared to the ABS CPB scheduling schemes. Moving towards more practical situations, we intend to consider imperfect CSI estimation for such systems in the future work.

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