On Performance of STAR-RIS-Enabled Multiple Two-Way Full-Duplex D2D Communication Systems

TIEN HOA NGUYEN1 AND TIEN TUNG NGUYEN2

1School of Electrical and Electronic Engineering, Hanoi University of Science and Technology, Hanoi 100000, Vietnam
2Faculty of Electronics Technology, Industrial University of Ho Chi Minh, Ho Chi Minh City 700000, Vietnam

Corresponding author: Tien Hoa Nguyen (hoa.nguyentien@hust.edu.vn)

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ABSTRACT

This paper investigates performance of simultaneously transmitting and reflecting reconfigurable intelligent surface (STAR-RIS)-enabled multiple two-way full-duplex device-to-device (D2D) communication systems over Rayleigh fading channels under optimal and uncertain phase shift alignments. We derive closed-form expressions for outage probability (OP), sum throughput, ergodic capacity (EC) and energy efficiency. To gain insights, we quantify and reveal some useful guidelines for the performance behavior of the OP and the EC, such as diversity order and ergodic slope from high transmit power configuration. In addition, some critical points also deduced for the sum throughput and the system energy efficiency. Moreover, the impacts of the transmit power configurations, RIS deployments, allocating target data rate transmission, and the number of user deployments on the system performance are examined. Finally, we present some extensive simulations using Monte-Carlo method to corroborate the accuracy of the theoretical analysis.

INDEX TERMS

Simultaneously transmitting and reflecting reconfigurable intelligent surfaces (STAR-RISs), device-to-device (D2D) communication, full-duplex, two-way communication.

I. INTRODUCTION

Recently, a new technological revolution, namely reconfigurable intelligent surfaces (RIS), has received widespread attention and is becoming a climactic solution to eradicate the limitations of spectral- and energy-efficiency (EE) for the six-generation (6G) of wireless communications [1]. Unfortunately, this RIS-based system only covers a half-space transmission while the terminals in practice may not be located on both sides of a RIS owing to the geographical restriction. Inspired by this fact, the authors of [2], [3] introduced a novel concept of simultaneously transmitting and reflecting (STAR)-RISs while [4], [5] released an intelligent Omni-surface (IOS). Although these works have the same idea to achieve full-space coverage, their architectural design is totally different. Specifically, the design of STAR-RIS mainly relies on the electric and magnetic currents of a STAR-RIS element to manipulate the transmitted and reflected signals. Meanwhile, the phase shifts for transmission and reflection in IOS systems are determined by the state of positive intrinsic negative (PIN) diodes installed in the IOS elements. In order to realize the concept for practice implementation, [6] invested a prototype for IOS while [7] and [8] proposed two prototypes using meta-surfaces to resemble STAR-RISs. Despite the above advantages, research into how STAR-RISs can be integrated into wireless communication systems is still in its infancy. Inspired by this, the research on STAR-RIS networks has been early investigated, e.g., improving security for multiple-input single-output network [9], and terahertz information and power transfer [10]. Very recent time, to exploit the full potential of STAR-RISs, the authors of [11] proposed three practical protocols:

• Energy splitting (ES) protocols: It typically consumes a relatively high overhead for configuration information exchange because of using large number of design variables.
• Time switching (TS) protocols: It is easy to handle the design of the transmission and reflection coefficients. However, this protocol introduces stringent requirements for time synchronization, raising high hardware implementation complexity.
• Mode switching (MS) protocols: It cannot achieve the same full-dimension transmission and reflection. However, this protocol does not require much energy assignment and has a simple hardware design when compared to ES and TS.

Owing to the explosion of 6G, an increase in the demand for data traffic will become essential. As a result, device-to-device (D2D) communications have become extremely crucial [12]. The combination of both RIS and D2D technologies in wireless systems have been considered. In [13], the authors showed that the interference of D2D networks can be significantly decreased by optimizing the transmit power and discrete phase shifts of RIS elements. A solution based on jointing power control of D2D users and passive beamforming for an RIS-aided D2D communication network for maximizing energy efficiency was proposed in [14]. In [15], a throughput analysis and the impact of the proposed mode selection mechanism of D2D communication of RIS-assisted D2D communication were investigated. Moreover, the authors analyzed to underlay and overlay modes of D2D communication. Considering imperfect hardware including both hardware impairment at the transceivers and phase noise at the RISs, the authors optimized the phase shift to maximize the achievable rate for both continuous phase shifts and discrete phase shifts in [16]. With goal of maximizing the overall spectrum efficiency and energy efficiency of the network, the resource reuse indicators, the transmit power and the RIS’s passive beamforming were optimized in [17]. Later, as an extent version of [17], the authors in [18] used an alternating algorithm integrated with Dinkelbach’s method to maximize he overall spectrum efficiency and energy efficiency. Recently, for maximizing the sum of the transmission rate of the D2D system and the cellular network, a novel solution based on decentralized double deep Q-network framework was proposed in [19].

With the emergence of the STAR-RIS architecture, D2D communication can contribute to more efficient information transmission, such as improved connectivity and significant cost savings [20]. However, there are a few shortcomings to STAR-RIS-enabled D2D communications [21]. On the other hand, with recent signal processing breakthroughs in self-and loop-interference cancellations, two-way full-duplex D2D (TWFD) communications have also emerged as a potential avenue for further performance improvement by sharing channels. However, to the best of our knowledge, no previous work analyzed the performance of multiple TWFD communications with STAR-RIS networks. Most of the existing works only focused on conventional RIS systems with either one-way communications [22], single two-way communications [23], [24], [25], or multiple two-way communications over orthogonal frequency-division-multiplexing (OFDM) [26].

Aiming to fulfill this gap, in this paper, we focus on analyzing the performance of STAR-RIS-based systems with multiple TWFD, where the base station plays to maintain and control the communication between pairing devices in the clusters via a RIS. In [26], applying multi-pair OFDM typically demands large frequency bands, and thus it is limited to the system with multiple devices and restricted frequency allocation. Therefore, in order to deal with such problems as well as fully exploit the inherent advantages of STAR-RISs, we consider the RIS operating in MS, where the whole number of elements of RIS is divided into multiple sub-surfaces (i.e., the number of sub-surface is identified by the number of pairing devices), and each sub-surface only services one paired D2D communication. It is worth noting that RIS with MS can readily be set up and lower overhead rather than using ES and TS and employing such RIS architecture with optimal phase-shift alignment (OA), i.e., configuring the beam of signal reflection to the destination, can reduce the impact of the co-channel interference from other sub-surfaces. Aside from that, by adopting some sophisticated interference cancellations in conventional full-duplex communication (e.g., successful interference cancellation - SIC) [27], employing TWFD devices becomes a reasonable solution and contributes significantly to the enhancement of spectral efficiency. However, it is impossible to completely eliminate the loop interference for full-duplex communication systems, and definitely occurs the residual self-loop interference. To deeply exemplify the benefits obtained from the considered system for future practical implementations, it is necessary to investigate the performance comparison between STAR-RIS with OA and uncertain phase shift alignments (UA) in the presence of residual loop interference under an in-avoidant co-channel one. Note that the considered system can be extended to various scenarios, such as heterogeneous networks, industrial networks (e.g., access point manages sensors communicating with others), overlaying cognitive networks (e.g., multiple transmitters and receivers communicate with each other via the same RIS), vehicle-to-everything communications, and so on. In general, the contributions of this work can be summarized as follows:

1) We investigate multiple TWFD-aided STAR-RIS systems over Rayleigh fading channels under two scenarios of optimal and uncertain phase-shift alignments.
2) We derive the closed-form expressions for the outage probability (OP), sum throughput, effective ergodic capacity (EC), and the EE.
3) We analyze the OP and EC at high power configurations to provide useful insights into system designs, such as diversity order and ergodic slope. Additionally, some critical observations for system throughput and EE are also represented.
4) We provide extensive simulation results to corroborate our derived expressions, and the results reveal
that: (i) varying transmit power configurations results in different outage performance, which can be either immensely improved or become saturate, (ii) increasing the RIS elements can only strengthened dramatically with OA, (iii) the throughput with OA achieves superior improvement over UA when raising transmission data rate, and (iv) when increasing the number of user, the EESC only increases significant with OA.

The rest of the paper is constructed as follows: Sec. II presents the system model, Sec. III provides theoretical analysis, Sec. IV shows numerical results and discussions, and finally Sec. V deduced the insights and obtained analysis.

**Notations:** $\Pr[-]$ denotes the expectation operator, $f_X(x)$ and $F_X(x)$ stand for the probability density function (PDF) and the cumulative distribution function (CDF) of a random variable $X$, respectively. $\Pr[-]$ indicates the probability function, $\mathcal{CN}(0, \sigma^2)$ implies the complex Gaussian noise with zero-mean and variance $\sigma^2$, and $\mathcal{N}(0, \sigma^2)$ means the normal distribution with zero-mean and variance of $\sigma^2$. $\Gamma(\cdot)$ and $\Gamma^*(\cdot, \cdot)$, and $\Gamma^*(\cdot, \cdot)$ denote the Gamma function and the lower and upper incomplete Gamma function [28, Eqs. (8.350.1) and (8.350.2)], respectively, while $Ei(\cdot)$ presents the exponential integral function [28, Eq. (8.211.1)].

**II. SYSTEM MODEL**

**A. SYSTEM DESCRIPTION**

Consider a STAR-RIS-enabled TWFD communication system consisting of a STAR-RIS having $2N$ units (i.e., $R$), a set of $K$-two antennas user (i.e., $U$), and a set of $K$-two antennas user (i.e., $D$), as shown in Fig. 1. In the considered system, we only consider $U$ and $D$ located in a specific zone of transmission or reflection, and each $U_k$ exchanges the information symbol with its paired $D_l$ via the STAR-RIS, which acts in mode switching to simultaneously satisfy all D2D communications, with $\{k, l\} \in \mathcal{K}$ and $\mathcal{K} = \{1, 2, \ldots, K\}$. Specifically, all STAR-RIS units assigned for the transmission or reflection zones are then divided into multiple sub-surfaces having $\mathcal{T} = N/K$ units and each sub-surface only serves for one pair ($U_k, D_l$).

![FIGURE 1. Illustration of system models.](image)

In this paper, we assume that: (i) there is no direct link between $U_k$ and $D_l$ owing to the face of obstacles and/or deep shadowing; (ii) all users use the same codebook, where $s_k$ and $c_l$ stand for the unit-energy information symbols from $U_k$ and $D_l$, respectively; (iii) all wireless links are assumed to be subject to Rayleigh fading distribution, and they are independent, no correlation exists, and each pair ($U_k, D_l$) can perfectly know the channel state information of each other; (iv) a pair of antennas at $U_k$ and $D_l$ are only responsible for transmission and reception; (v) the pairing process between the devices and configures the STAR-RIS operation (i.e., selective frequency, calculates the phase shift, aligns the phase shift, and sets up the STAR-RIS units according to the specific communication) are controlled and managed directly by a base station; and (vi) the STAR-RIS design is linearly parametrized, i.e., all inter-element at the STAR-RIS has spaced more than half a wavelength apart, and thus, the impact of mutual-coupling information/joint processing over the STAR-RIS element can be ignored [31].

**B. CHANNEL MODELS**

Considering the far-field regions, the wireless channels of $U_k \rightarrow R$ and $R \rightarrow D_l$ links are able to be regarded as reciprocal channels. Thus, the forward and backward channels between $U_k$ (or $D_l$) and $R$ are the same, shown in Fig. 1. In this sense, we denote $\mathbf{h}_k$, $\mathbf{g}_l$, $\hat{\mathbf{h}}_k$, and $\hat{\mathbf{g}}_l$ be the channel vectors of $U_k \rightarrow R$ links, the channel vectors of $R \rightarrow D_l$ links, channel loop interference of $U_k$, and channel loop interference of $D_l$, respectively. Since wireless channel fading complies with Rayleigh distribution, each entry input of $\mathbf{h}_k$ (i.e., $h_{k,n}$) and $\mathbf{g}_l$ (i.e., $g_{l,n}$), $\hat{\mathbf{h}}_k$, and $\hat{\mathbf{g}}_l$ can be modeled as complex Gaussian fading with zero-mean and the corresponding variances of $\lambda_k$, $\lambda_l$, $\hat{\lambda}_k$, and $\hat{\lambda}_l$. Also, the diagonal phase shift matrix introduced by the STAR-RIS is denoted as $\Phi_{k,l} = \text{diag}(\varphi_{k,0}, \varphi_{k,1}, \ldots, \varphi_{k,K-1})$, where $\varphi_{k,n} \in [-\pi, \pi]$ and $n \in \mathcal{N}$, with $\mathcal{N} = \{0, 1, 2, \ldots, K\}$. Finally, for convenient presentation, we denote $\{p, q\} \in \mathcal{K}$ such that $q \neq k$ and $p \neq l$, $\mathcal{M} = \{(q - 1)\mathcal{I} + 1, \ldots, q\mathcal{I}\}$, and $m \in \mathcal{M}$.

**C. INFORMATION TRANSMISSION PROCESS**

In the communication, since each sub-surface is assigned a phase shift profile to the corresponding paired D2D interface and all the sub-surfaces operate simultaneously, each user ($U_k$ or $D_l$) receives not only a superposition of two signals by its sub-surface but also suffers the co-channel interference from other surfaces. Similar to [22], we only focus on the first signal arrival to the receiver, while those reflected by the STAR-RIS two and more times are omitted due to the substantial path loss in the far-field environment. For simplicity, we first examine the received signal at $U_k$ at the time $t$-th
as follows:

$$\gamma_{U[k]} = \sqrt{P_{D_L} \left( \sum_{n} h_{k,n} e^{j\phi_{k,n}^a} g_{l,n} \right)^2}$$

where $w_k \sim \mathcal{CN}(0, \sigma^2)$, and $P_U$ and $P_D$ stand for the transmit power of the set of $U$ and $D$, respectively. Since $U_k$ has the knowledge of $e^{j\phi_{k,n}^a}$, $h_{k,n}$, $g_{l,n}$, and $s_k$, the self-interference term can readily eliminate. Meanwhile, in order to reduce the impact of loop interference, $U_k$ is assumed to be able to adopt some sophisticated interference cancellations similar to conventional full-duplex communications [23]. However, residual self-loop interference is unavoidable. Therefore, to better characterize the impact of residual self-loop interference for practical applications, we use the modeling of $\lambda_k$ similar to [27], where $\lambda_k = \omega P_{W_U}$ with $\omega > 0$ and $\eta \in [0, 1]$ being two constant values. After the elimination, the received instantaneous signal-interference-plus-noise ratio (SINR) at $U_k$ can be written as

$$\gamma_{U[k]} = \frac{\sum_{n} h_{k,n} e^{j\phi_{k,n}^a} g_{l,n}^2}{\sum_{q,p,m} h_{q,m} e^{j\phi_{q,m}^a} g_{p,m}^2},$$

where $P_{Uk}$, $P_{Dk}$, and $P_{Dk}$ present the transmit powers of users $U_k$, $D_k$, and $D_q$, respectively. The instantaneous SINR at $V_{\theta}$ is $\gamma_{V_{\theta}}$, with $\theta \in \{k, l\}$, is written as

$$\gamma_{V_{\theta}} = \frac{\sum_{n} h_{k,n} e^{j\phi_{k,n}^a} g_{l,n}^2}{\sum_{q,p,m} h_{q,m} e^{j\phi_{q,m}^a} g_{p,m}^2} + \sigma^2 + \lambda_{D_q},$$

where $S_{\theta} \subseteq \{U_k, D_k\}$ but $S_{\theta} \neq V_{\theta}$.

III. PERFORMANCE ANALYSIS

A. OPTIMUM PHASE-SHIFT ALIGNMENT

With the goal of maximizing the instantaneous SINR of $(U_k, D_k)$, the phase-shift $\phi_{k,n}$ in (3) should be configured in the form $\phi_{k,n} = -\angle k_{n}^a - \angle g_{l,n}$ [31]. This setup is possible at the STAR-RIS due to the achieving global phase infusion of the respective channels. As such, $\sum_{n} h_{k,n} e^{j\phi_{k,n}^a} g_{l,n}$ can be rewritten as $\chi_1 \triangleq \sum_{n} h_{k,n} |g_{l,n}|$. Since $|h_{k,n}|$ and $|g_{l,n}|$ follow Rayleigh distributions, the first and second moments of $\chi_1$ can be calculated as $\mathbb{E}[\chi_1] = \mathbb{T} \pi / 4$ and $\text{Var}[\chi_1] = (1 - \pi^2 / 4) \lambda_k^a / 4$, respectively. By fitting the shape parameter $\alpha = (\mathbb{E}[\chi_1]^2 / \text{Var}[\chi_1]) = \pi^2 / (16 - \pi^2)$ and the scale parameter $\beta_1 = \text{Var}[\chi_1] / \mathbb{E}[\chi_1] = (4/\pi - \pi/4) \lambda_k / 4$, we obtain the CDF of $\chi_1$ in the form of Gamma distribution, i.e.,

$$F_{\chi_1}(x) = \frac{x^{(\alpha - 1)} e^{-(x/\beta_1)}}{\Gamma(\alpha)}.$$  (4)

By using the fact that $F_{\chi_1}(x) = F_{\chi_2}(x)$, we obtain the CDF of $\chi_2$ in closed-form expression as

$$F_{\chi_2}(x) = \frac{x^{(\alpha - 1)} e^{-(x/\beta_1)}}{\Gamma(\alpha)} - 1.$$  (5)

By using the distribution of $\phi_{q,m}$ in [32], we have

$$\mathbb{E}[\mathbb{R}(\chi_2)] = \mathbb{E}[\mathbb{R}(\mathbb{Z})] = 0, \quad \text{Var}[\mathbb{R}(\mathbb{Z})] = \frac{\lambda_1}{2}. \quad (6)$$

Although the channels from $p$-th $D$ to $q$-th $U$ over $q$-th sub-surface of RIs can be regarded as the same distribution, the summation of the channels from $p$-th $D$ to $q$-th $U$ are typically different owing to varying location. Thus, $\chi_2$ has an independent and non-identical distributed (i.n.i.d). Thus, the first and second moment of $\chi_2$ can be determined as

$$\mathbb{E}[\mathbb{R}(\chi_2)] = \mathbb{E}[\mathbb{R}(\chi_2)] = 0, \quad \text{Var}[\mathbb{R}(\mathbb{Z})] = \frac{\lambda_1}{2}. \quad (9)$$

From (7) and (9), we can express $\mathbb{E}[\mathbb{R}(\chi_2)] = \mathbb{E}[\mathbb{R}(\chi_2)] = 0$ and $\text{Var}[\mathbb{R}(\mathbb{Z})] = \frac{\lambda_1}{2}$. On the other hand, $|\chi_2|^2$ can be written as $|\chi_2|^2 = |\mathbb{R}(\chi_2)|^2 + |\mathbb{I}(\chi_2)|^2$. Using moment matching, we have

$$\mathbb{E}[|\chi_2|^2] = \text{Var}[|\chi_2|^2] + (\mathbb{E}[|\chi_2|^2])^2 = \sigma_{re}^2 + \sigma_{im}^2. \quad (11)$$

$$\text{Var}[|\chi_2|^2] = \text{Var}[|\chi_2|^2] - (\mathbb{E}[|\chi_2|^2])^2 = 2(\sigma_{re}^4 + \sigma_{im}^4). \quad (12)$$

Now, we use the moment-matching method to fit the CDF of $|\chi_2|^2$ with Gamma distribution [31], with the shape $\alpha_2 = (\mathbb{E}[|\chi_2|^2] / \text{Var}[|\chi_2|^2]) = 1$ and the scale $\beta_2 = \text{Var}[|\chi_2|^2] / \mathbb{E}[|\chi_2|^2] = \sigma_{re}^2 + \sigma_{im}^2$. Therefore, $|\chi_2|^2 = \chi_2$.}

Lemma 1: The PDF of $|\chi_2|^2$ can be rewritten as

$$f_{|\chi_2|^2}(x) = \exp(-x/2) / \beta_2. \quad (13)$$
1) OUTAGE PROBABILITY

By definition, the outage probability of $V_\theta$ is expressed as $OP_{V_\theta} = \Pr[y_\theta \leq y_{th}]$, where $y_{th} = 2^{r_0}$ is the SINR threshold while $r_0$ is the target data rate.

From (3), (5), and (13), we have

$$OP_{V_\theta} = \int_0^\infty \frac{\gamma_{th}(\theta)}{S_\theta} f_{|x|^2}(y) dy = \int_0^\infty \frac{\gamma_{th}(\theta)}{\beta_1} \exp\left(-\frac{\gamma_{th}}{\beta_1}\right) d\gamma_{th},$$

(14)

where $\Delta_1 = [P_{D_q} + P_{U_p}] / P_{S_\theta}$ and $\Delta_2 = (\sigma^2 + \tilde{\sigma}_o) / [P_{D_q} + P_{U_p}]$.

Although the integral in (14) can solve by using the expansion series in [28, Eq. 8.354] and the help of the identity [28, Eq. (3.382.4)], the obtained closed-form expression is an infinite series with low convergence, i.e.,

$$OP_{V_\theta} = \sum_{\gamma_{th} = 0}^{\infty} \frac{\gamma_{th}(\theta)}{\beta_1} \exp\left(-\frac{\gamma_{th}}{\beta_1}\right),$$

(15)

To deal with such a mathematical problem, we next adopt an efficient approximation method with high accuracy.

**Proposition 1:** Invoking (5) and performing the transformation of $z = y/\beta_1$, (14) can approximate in closed-form thank to the help of generalized Gauss-Laguerre quadrature as

$$OP_{V_\theta} \simeq 1 - \sum_{i=1}^{J} \xi_i \Gamma\left(\alpha_1, \gamma_{th}(\theta) \Delta_1/\beta_1\right),$$

(16)

where $J$ stands for the accuracy-complexity trade-off coefficient, $\Delta_i$ presents the $i$th root of the generalized Laguerre Polynomial provided by

$$L_0^{\alpha}(\delta) = \sum_{j=0}^{J} \frac{(\delta + 1)^j}{j!} g_j,$$

(17)

and $\xi_i$ implies the weight vector that is expressed as

$$\xi_i = \delta_i / [J + 1L_0^{\alpha}(\delta_i)].$$

(18)

**Remark 1:** Expression in (16) shows some important observations: (1) when $P_{U_\theta} \in \{P_{D_q}, U_{p}, V_\theta\}$, is fixed (i.e., $\Delta_2$ is always constant), raising $P_{S_\theta}$ can reduce $OP_{V_\theta}$, yielding $\Delta_1 \to 0$ and $OP_{V_\theta} \propto 1/P_{S_\theta}^{\alpha_1/2}$. Thus, one can deduce that the diversity order of $V_\theta$ is $\alpha_1/2$.

(2) when $P_{S_\theta} = P_{D_q} = P_{U_p}$, increasing $P_{S_\theta}$ results in $\Delta_1 \to 2$ and $\Delta_2 \to 0$. Thus, there exists an outage floor for $V_\theta$, which means that $V_\theta$ has zero-diversity order.

(3) when $P_{S_\theta} = P_{D_q}(P_{U_p})$, $P_{V_\theta}$ tends to decrease and then converges on outage floor, resulting in zero diversity order for $V_\theta$. (4) when $P_{S_\theta}$ and $P_{D_q}$ are fixed, increasing $P_{V_\theta}$ occurs two important cases: (i) $P_{V_\theta}$ is larger than $P_{U_p}$ or $P_{D_q}$ (i.e., the nodes are two same transmitting side with $V_\theta$) and (ii) $P_{U_p} = P_{U_p}$ or $P_{V_\theta} = P_{D_q}$. For the first case, the OP of $P_{V_\theta}$ has very slight change because $\tilde{\sigma}_o = \omega P_{V_\theta}$ is relatively small, with $\omega \in (10^{-4}, 10^{-3})$ [23]. For the second case, the OP of $P_{V_\theta}$ increases significantly with the increase of $P_{V_\theta}$, i.e., $\Delta_1 \to \infty$.

2) SUM THROUGHPUT

Throughput in delay limited-mode transmission is defined as the amount of information data that can be successfully delivered over fading channels when given a constant rate $r_0$, and it can be expressed mathematically as $r_\theta = (1 - P_{V_\theta}^\text{out}) r_0$.

The sum throughput of the considered system is

$$r_{\text{sum}} = \sum_{i=1}^{2k} r_i = \sum_{i=1}^{2k} (1 - P_{V_i}^\text{out}) r_i,$$

(19)

**Remark 2:** As can be seen, $r_{\text{sum}}$ is not only dominated by $r_\theta$ but also $P_{V_i}^\text{out}$. However, since there are several situations related to Remark 1, the achieved $P_{V_i}^\text{out}$ is also varying. In such cases, dynamically assigning $r_\theta$ for its corresponding user becomes very important, posing an interesting mathematical throughput optimization problem.

3) EFFECTIVE ERGODIC CAPACITY

According to Shannon theorem, the capacity of the system over fading channel is defined as $C = \log_2(1 + \text{SINR})$ [bit/sec/Hz] [33], [34]. Thus, the ergodic capacity of $V_\theta$ in the considered system can be evaluated as $C_\theta = \int_0^\infty \log_2(1 + y_{\theta}) f_{y_{\theta}}(y) dy$. Using partial integration transformation, $C_\theta$ is rewritten as

$$C_\theta = \frac{1}{\ln 2} \int_0^\infty \frac{1 - F_{y_{\theta}}(x)}{1 + x} dx,$$

(20)

From the facts that $P_{V_i}^\text{out} = F_{y_{\theta}}(x)$, we get

$$C_\theta \simeq \frac{1}{\ln 2} \sum_{i=1}^{J} \frac{\xi_i}{\Gamma(\alpha_1)} \int_0^\infty \frac{1}{1 + x} \Gamma\left(\alpha_1, \sqrt{x} \Delta_3 / \beta_1\right) dx,$$

(21)

where $\Delta_3 \triangleq \sqrt{\Delta_1(\beta_2 \delta_1 + \Delta_2)}$. By transforming the incomplete Gamma function to the equivalent Meijer-G function, we obtain

$$\Gamma\left(\alpha_1, \sqrt{x} \Delta_3 / \beta_1\right) = G_{1,2}^{0,1}\left(\sqrt{x} \Delta_3 / \beta_1\right)^{1/0, \alpha_1}.$$

Following that, (21) becomes

$$C_\theta \simeq \frac{1}{\ln 2} \sum_{i=1}^{J} \frac{\xi_i}{\Gamma(\alpha_1)} \int_0^\infty \frac{1}{1 + x} G_{1,2}^{2,0}\left(\sqrt{x} \Delta_3 / \beta_1\right)^{0, \alpha_1} dx,$$

(22)

where the integral in (22) is obtained thanks to the help of [28, Eq. (7.811.5)]. Based on the analysis above, the effective ergodic sum capacity (EESC) of the system is deduced as

$$C_{\text{sum}} \simeq \sum_{j=1}^{2k} C_j,$$

(23)
By the virtue of asymptotic analysis, we provide the ergodic slope to capture the influence of channel parameters on the EESC. The ergodic slope is computed as [23]

\[
S_{\phi} = \lim_{P_{V_{\phi}} \to \infty} \frac{C_{\text{sum}}(P_{V_{\phi}})}{\log(P_{V_{\phi}})} = 0, \quad (24)
\]

\[
S_{\phi} = \begin{cases} 
1, & P_{S_{\phi}} > \max P_{l|3} | P_{l|3} = c, \\
0, & P_{S_{\phi}} = \max P_{l|3} | P_{V_{\phi}} = c,
\end{cases} \quad (25)
\]

where \(c\) is constant value.

Remark 3: The above analyses show that: when \(V_{\phi}\) increases its transmit power, i.e., \(P_{V_{\phi}} \to \infty\), the EESC is degraded. However, when increasing \(P_{S_{\phi}}\), the EESC tends to increase with the ergodic slope of 1 if \(P_{S_{\phi}} > \max P_{l|3}\) \(\phi\), otherwise, it converges the throughput ceiling and achieves zero ergodic slopes even when there is no interference signal between the antennas of the users.

4) ENERGY EFFICIENCY

To provide quantitative analysis for practical implementations, we further investigate the EE, which is defined as the ratio between the total data rate of all users and the total energy consumption. Based on the analysis in (19) and (23), the EE of the considered system is given by

\[
\mu_{EE}^{x} = \frac{\frac{1}{T} \sum_{k} P_{U_{k}} + \sum_{i} P_{D_{i}} + P_{C}}{\log(1 + P_{V_{\phi}})}, \quad (26)
\]

where \(x \in \{ t_{\text{sum}}, c_{\text{sum}} \}\), \(P_{C}\) denotes the total circuit consumption, and \(T\) is transmission duration. Of note, \(t_{\text{sum}}^{x}\) implies the system EE in delay-limited transmission mode (DLT) while \(c_{\text{sum}}^{x}\) indicates the delay-tolerant transmission one (DTT).

Remark 4: The expression in (26) shows that \(P_{U_{k}}\) and \(P_{D_{i}}\) not only affect directly to the \(x\) (shown in Remarks 2 and 3) but also decides on the \(\mu_{EE}^{x}\) of the system. Thus, it also raises another question on how to design a reasonable transmit power policy in TWFD-aided STAR-RIS systems.

B. UNCERTAIN (ERROR) PHASE-SHIFT ALIGNMENT

In practice, the phase-shift configuration at the RIS is able to occur errors due to the failure of channel estimation or phase discretization. As such, \(\phi_{k,n}\) can be regarded as the phase adjustment error \(\phi_{\text{err}}\), which is subject to uniform distributions \(U(-\pi, \pi)\) [23]. Therewith, we obtain the the PDF of \(|x_{1}|^{2}\) in closed-form expression (similar to derive \(|x_{2}|^{2}\)) as

\[
f_{|x_{1}|^{2}}(x) = \exp(-x/\hat{\beta}_{1})/\hat{\beta}_{1}, \quad \hat{\beta}_{1} = 1, \quad \text{for} \quad |x_{2}|^{2} \text{as}
\]

\[
f_{|x_{1}|^{2}}(x) = \exp(-x/\hat{\beta}_{1})/\hat{\beta}_{1}, \quad \hat{\beta}_{1} = 1, \quad \text{for} \quad |x_{2}|^{2}
\]

Following that, we attain the OP for \(V_{\phi}\) similar to (14) as

\[
P_{V_{\phi}}^{\text{out}} = 1 - \int_{0}^{\infty} \int_{0}^{\infty} f_{|x_{1}|^{2}}(x)f_{|x_{2}|^{2}}(y)dx dy = 1 - \frac{\exp(-\gamma_{h}^{\Delta_{1}}/\hat{\beta}_{1})}{\hat{\beta}_{1}^{\Delta_{1}}/\hat{\beta}_{1}}, \quad (28)
\]

By replacing (28) into (20), one can get

\[
C_{\phi} = \frac{1}{\ln 2} \int_{0}^{\infty} \frac{\exp(-x/\hat{\beta}_{1})}{(1 + x)^{\hat{\beta}_{1}}/\hat{\beta}_{1}} dx
\]

\[
= \frac{1}{\ln 2} \left[ \frac{\hat{\beta}_{1}}{\hat{\beta}_{1}} \left[ Ei\left(-\frac{\Delta_{1}}{\hat{\beta}_{1}}\right) - Ei\left(-\frac{\Delta_{1}}{\hat{\beta}_{1}}\right) \right] \right], \quad (29)
\]

where (29) can be obtained by employing the concept of partial fractions and then using the aid of [28, Eq. (3.354.4)].

Finally, by inserting \(P_{V_{\phi}}^{\text{out}}\) and \(C_{\phi}\) into (19) and (23), respectively, we obtain the sum throughput and EESC, based on this, the system EE is also deduced.

IV. NUMERICAL RESULTS AND DISCUSSIONS

This section provides extensive simulation using the Monte-Carlo method to validate the analysis in Sec. III. In particular, all the channels are considered with free space path-loss propagation, i.e., \(\lambda_{k} = C_{d_{k}}^{-1}\) and \(\lambda_{I} = C_{d_{I}}^{-1}\). Herein, \(C_{d_{k}}\) and \(C_{d_{I}}\) stand for the physical distance in meter between \(U_{k}\) and \(R\) and between \(D_{I}\) and \(R\), respectively, \(L\) represents the attenuation power at 20 dB, and \(\mu\) is the path-loss exponent. Unless other specific, we set \(K = 4, r_{j} = 1, \mu = 3, \text{power noise} \sigma^{2} = 10^{-4}\) and \(\eta = 0.5, 10^{6}\) channel realizations.

Fig. 2 shows the OP comparison of \(V_{\phi}\) between OA and UA. Several observations are obtained: (i) the theoretical analyses (i.e., the curves with continuous and dash-lines) in (16) and (28) agreeably match with the simulation results (i.e., markers), corroborating the accuracy of our analysis. (ii) For the different locations of paired D2D users and fixed \(P_{l|3}\) at 5 dBm in Fig. 2(a), we notice that the OP of \(V_{\phi}\) linearly decreases as \(P_{S_{\phi}}\) increases and reduces significantly when paired D2D users are located near the RIS and descend at their distal location. (iii) By configuring \(C_{d_{k}} = 10, C_{d_{I}} = 15\), and \(P_{V_{\phi}} = 5\) dBm in Fig. 2(b), the performance of \(V_{\phi}\) with scheme 1 (i.e., \(P_{S_{\phi}} = P_{D_{k}}(P_{U_{k}})\)) is worse than that with scheme 2 (i.e., \(P_{S_{\phi}} = P_{D_{k}}(P_{U_{p}})\)) at low power settings but outperforming at moderate and high power ones. As such, for the scheme 1, it is crucial to increase the transmit power from \(S_{\phi}\), while for the scheme 2, it is suggested to increase the number of the reflective elements of the RISs, i.e., \(N\), and/or adjust target data rate transmission to improve the performance of \(V_{\phi}\). (iii) the performance of \(V_{\phi}\) with OA is outstanding than that of UA in both Figs. 2(a) and (b) and it is rashly degraded with UA in Fig. 2(b) due to strong co-channel interference.

Next, in Fig. 3, we compare the OP between the considered system and three relaying benchmark schemes using orthogonal frequency, i.e., amplifier-and-forward full-duplex (AFFD), decode-and-forward half-duplex (DFHD), and decode-and-forward full-duplex (DFFD), and all paired devices has the same sub-bandwidth (i.e., the original bandwidth is divide into \(K\)-sub-bandwidth). For full-duplex
T. H. Nguyen, T. T. Nguyen: On Performance of STAR-RIS-Enabled Multiple Two-Way Full-Duplex Communication, the modeling of loop interference at the devices and relay is assumed to be the same. As can be observed, the OP of the considered system with UA suffers significant performance degradation when compared to the benchmark schemes. The main reason for this phenomenon can be explained as follows: (1) error the phase-shift leads to varying reflective signals and thus the end-to-end (e2e) signals do not enhance. (2) The number of RIS elements assigned for each subsurface is relatively small, e.g., $\mathcal{I} = \{100/K = 25, 200/K = 50, 300/K = 75\}$. (3) strong co-channel interference caused by other device signals, which can be found in the first denominator of equation (3). However, employing such benchmark schemes requires at least $K$ orthogonal frequency bands and $K$ relays or one relay-equipped with $K$-pair antennas. This setup not only consumes a large number of frequencies (i.e., narrowing the bandwidth used for each paired device) but also increases the undesired overhead for practical implementation. Regarding the case of OA, when increasing the number of RIS units and transmitting power, the OP of the considered system improves significantly. This is because optimizing phase-shift design leads to maximizing e2e signal received at the devices, and thus the received signal of each device becomes much better than that of UA.

In Fig. 4, we examine the effect of $N$ and $r_{\phi}$ on the sum throughput of one set $V_{\phi}$ (i.e., either $U$ or $D$) for the scheme 1 and $(d_k, d_l)$ is set as $\{(10, 20), (25, 15), (5, 30), (15, 5)\}$. As can be seen in Fig. 4(a), the sum throughput with OA increases considerably with an increment of $N$ but does not change with UA. This is due to the fact that in the case of OA, allocating higher $N$ at RIS contributes a better signal for the user, while for the UA, the intensity of desired signals is much lower.
lower than the total other ones. Aside from that, it can be clearly seen that the sum throughput with OA outperforms the benchmark schemes in most of transmit power ranges. This is because the transmission rate per sub-bandwidth is less than $K$ times when compared to that of the considered system. Meanwhile, for the worse case, i.e., UA, the considered system achieve the approximate sum throughput when compared to the DFFD scheme and better than that of AFFD and DFHD. Next, we inspect the sum throughput under varying $r_\theta$, shown in Fig. 4(b). As can be observed, when $r_\theta$ increases from 0.5 to 1, the sum throughput with OA increases ghastly from 1.5 to 4 bit/s/Hz, while for the UA, it only increases slightly from 0.5 to 1 bit/s/Hz. Note that the performance increment of OA is five times that of UA.

Continuously, we explore the impact of $P_{V_\theta}$ and the number of users on the EESC of one set $V_\theta$ in Fig. 5. As can be observed in Fig. 5(a), when $P_{S_\theta}$ increases, the EESC curves tend to increase and then reach saturation. Besides, the figure also showed that EESC with OA provides better performance than UA. However, its performance drops more sharply than UA when $P_{V_\theta}$ increases from 5 dBm to 15 dBm. Fortunately, when configuring $P_{S_\theta}$ among (20,30) dBm, this phenomenon can be well solved, bringing significant performance gap improvements in the case of OA. Besides, compared to benchmark schemes, the EESC with OA achieves the best performance, while for the UA, it is only better in moderate and high transmit power regimes. Fig. 5(b), we plot the EESC with change of $K$. It is observed that when $K$ increases from 2 to 6, the EESC with OA increases rapidly while UA increases very slowly. Hence, one conclusion can be deduced that multiple TWFD communications can only work well with the RIS implementation if and only if the OA is carried out.

Finally, in Fig. 6, we plot the EE of the considered system in (26) with the same distance setting $(d_k, d_l)$ in Fig. 4. The first observation is that all the EE curves tend to decrease linearly with the increment of the transmit power. Also, when $N$ increases, the RIS with UA has almost the same EE in both DLT and DTT, whereas the EE of the RIS with OA tends to increase. Moreover, the EE in case of DTT outperforms that in case of DLT in most transmitting power regimes.

V. CONCLUSION

This paper has studied the performance of the TWFD-aided STAR-RIS system under two types of phase-shift alignment, i.e., OA and UA. The closed-form expressions for the OP, sum throughput, EESC, and EE have been derived. Considering high transmit power configurations, the diversity order and ergodic slope were quantified to gain some useful insights into system designs. Owing to varying transmit power setups, the outage performance and ergodic capacity could be operated in linear mode or in saturation. STAR-RIS with OA always achieves superior performance when compared to UA in terms of OP, throughput, EESC, and EE have been derived. Considering high transmit power configurations, the diversity order and ergodic slope were quantified to gain some useful insights into system designs. Owing to varying transmit power setups, the outage performance and ergodic capacity could be operated in linear mode or in saturation. STAR-RIS with OA always achieves superior performance when compared to UA in terms of OP, throughput, EESC, and EE. By increasing the RIS units and the transmission data rate, the throughput with OA is greatly improved. Moreover, elevating the number of user deployments also contributes to an increase in EESC in the case of OA.

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TIEN HOA NGUYEN received the Dipl.-Ing. degree in electronics and communication engineering from Hanover University. He has worked at the Research and Development Department of image processing and the Development of SDR-Based Drivers, Bosch, Germany. He devoted three years of experimentation with Mimoon’s Research and Development Team to develop embedded signal processing and radio modules for LTE-A/4G. He worked as a Senior Expert at Viettel IC Design Center (VIC) and VinSmart for development of advanced solutions for aggregating and splitting/steering traffic at the PDCP layer and above, to provide robust and QoS/QoE guaranteeing integration between heterogeneous link types in 5G systems. Currently, he is a Lecturer at the School of Electronics and Telecommunications, Hanoi University of Science and Technology. His research interests include resource allocation in B5G&6G, massive MIMO, and vehicular communication systems.

TIEN TUNG NGUYEN received the B.Sc. and M.Sc. degrees from the University of Science Ho Chi Minh City, Vietnam, in 2005 and 2010, respectively, and the Ph.D. degree in electronics from Myongji University, South Korea, in 2021. Since 2011, he has been a Lecturer at the Industrial University of Ho Chi Minh City. His research interests include emerging topics of wireless communication for 5G&6G, including energy harvesting, physical layer security, cognitive radio, non-orthogonal multiple access (NOMA), short-packet communications, the Internet of Things (IoT), and applications of optimization and machine learning for wireless communications.

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