RIS-Assisted Spatial Modulation and Space Shift Keying for Ambient Backscattering Communications

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Abstract—In wireless communications, reconfigurable intelligent surfaces (RIS) are emerging as a promising technology that is made possible by the advent of software controlled metamaterial sheets for controlling the wireless channels dynamically. In future applications, IoT devices will have small sizes and limited power supply. To make these devices spectrally and energy efficient in accordance with the advanced 5G and 6G specifications, we propose ambient backscattering (ABSc) technique along with spatial modulation (SM) and space shift keying (SSK) for data transfer assisted with RIS. We also conduct a thorough performance analysis of these schemes in terms of outage probability, and bit error rate, validated by Monte-Carlo simulations, and provide comparative results that illustrate the merits of the proposed techniques. In particular, it is shown that RIS-empowered SM and SSK along with ABSc perform much better than conventional communications.

Index Terms—RIS, ABSc, SM, SSK, BER, Outage.

I. INTRODUCTION

It is expected that by 2030, the demand for data rates will reach 1 Tb/s due to the continuing increase in data traffic and wireless connections, which will be almost 1000 times more than what the 5G networks will provide. To meet these demands, the future 6G communication aims to attain spectral efficiencies of around 100 bps/Hz, with ultra-low latency of less than 1 ms. Such requirements are indeed key to support the envisioned applications such as extended reality services, smart healthcare, automated vehicle control, cellular connectivity, etc. Smart systems in 6G will also require small devices capable of interacting with each other without utilizing the base station resources, thereby saving power and cost, and eliminating power hungry transceivers due to the device size constraints. In this vein, alternative methods which can support the functioning of such smart devices in IoT networks, while adapting to the dynamic nature of the wireless channels, need to be developed. This makes it necessary to control and manipulate the radio waves so as to steer the radiation beams in the desired directions.

For such dynamic control, software controlled metasurfaces have come to the forefront, where the reconfigurability of RIS structures is ensured by the use of PIN diodes or varactors [1]. This allows adaptation to the dynamics of the wireless channels, and generation of multiple beams in the optimized directions. Cross-polarized and split ring resonator antennas can be used to design a RIS, where PIN diodes are used to make the resonators as reflectors or absorbers of the incident waves by connecting or disconnecting to the ground depending on the external voltage [2]. Varactor diodes and micro-electromechanical (MEMS) can be used to electronically change the resonant frequency and generate desired phase and amplitude changes for the reflected signal. Such components, present in a thin-film software-controlled RIS, can also exploit the multipath propagation, thus improving network coverage. Considering RIS as a perfect reflector, if a RF wave strikes a RIS, the received power gets scaled up by a factor equal to the square of the number of constituent units of RIS due to its scaling property [3]. The distance between the transmitter and RIS allows scaling of received power, enabling reduction of transmit power and enhancement in data rates. In the literature, RIS has been used as an access point and/or as a reflector. In [3] and [4], symbol error rate upper-bounds for RIS-assisted communication were provided. The upper bounds are quite tight when the number of RIS elements is high, but not so accurate for low numbers of elements. In [5], RIS-assisted non-orthogonal multiple access was analyzed in terms of bit error rate (BER).

Apart from the control of RF waves, energy harvesting for data transfer is another prime technology for future wireless communications, especially for low-power IoT applications. In particular, ambient backscattering (ABSc) is a promising technique by means of which devices can harvest energy from ambient RF waves, e.g., TV towers, WiFi devices or RF interference, and generate the required power for data transfer from tags to readers [2], [6]. In the transmit device, known as tag, antenna impedance variation changes the reflection coefficient, as a result of which the RF waves from nearby WiFi device(s) can be reflected or absorbed by the tag. Another device called reader, located near the tag, can receive the data by demodulating the wave which is reflected/backscattered by the tag. If the waves are absorbed by the tag, bit 0 will be transmitted, and if the waves are reflected, then bit 1 is sent. In ABSc communications, MIMO technology can also be used at the tag to achieve the required high spectral efficiency and high data rates. From a conceptual viewpoint, MIMO suffers from inter-antenna interference, unnecessary power consumption, and high cost of multiple RF chains. So far, spatial modulation (SM) has been shown to be notable in resolving these problems [7]. SM uses a single RF chain and...
activates one antenna within a transmission slot. In SM, a chunk of bits out of the information stream are considered at a time and split into two parts: \( \log_2(M) \) bits from the least significant bit (LSB) side are used for symbol mapping according to \( M \)-ary modulation, while the remaining \( \log_2(N_t) \) bits are used for single antenna activation, where \( N_t \) denotes the number of transmit antennas. Hence, the spectral efficiency (SE) of SM is \( \log_2(N_t) + \log_2 M \). Among several variants of SM, space shift keying (SSK) is a prominent one, where only the antenna index is used to convey the information [7]. The SE of SSK is given by \( \log_2(N_t) \). In the literature, SM has been applied in many MIMO communication models. In particular, [8] considered SM for ABSc-based applications, and provided upper bounds on the BER. In [9], SM and SSK techniques have been used at the receiving end using RIS as a reflector. Besides, spatial modulation to select particular radiation patterns of RIS and transmitter antennas simultaneously, have been performed in [10].

Looking at the potentiality of the above-mentioned technologies and motivated by the need for spectrally and energy efficient data transfer, we propose a RIS-assisted SM and SSK-based ABSc communication framework. Specifically, we propose tag devices capable of implementing ABSc to utilize the energy from ambient RF waves generated by WiFi routers. Tag devices carry out SM and/or SSK, and utilize RIS as a reflector to transfer data to reader devices. Using RIS allows us to increase the effective distance of the data transfer between tag and reader, and to reduce the transmit power requirement. Considering Nakagami fading model for the individual links, we derive upper-bound expressions for the BER and the outage probability for system operation with SM and SSK separately. Analysis of the power consumption of the tag is also demonstrated to highlight the effectiveness of the proposed techniques. Comparative results are provided, and illustrate the notable performance of the RIS-assisted SM and SSK with ABSc.

II. THE PROPOSED COMMUNICATION MODEL

A. RIS-Aided ABSc Communication with Spatial Modulation

The proposed communication model shown in Fig. 1 implements SM along with ABSc. The tag comprises an energy harvesting circuitry to harness power from WiFi signals generated by a single-antenna router. The bit mapper inside the tag splits the incoming bit stream for antenna selection and symbol mapping. For SSK, the bit mapper directly interacts with the antenna selector, with no symbol modulator. The antenna selector chooses a single antenna out of the available \( N_t \) antennas. The RIS comprises \( K \) elements, each of which is capable of introducing a phase shift to the incident signal coming from the selected antenna. The RIS can be designed by covering the walls of a room with a thin sheet of electronically tunable metasurfaces. The reader receives the signal reflected by the RIS. The reader is considered to have a single antenna to avoid complex detection circuitry and excessive power requirements, thereby justifying the benefits of ABSc communication.

For \( M \)-ary modulation using a tag, the modulating impedance values have to be chosen such that all the reflection coefficients are confined within a circle about the conjugate match having magnitude \( \leq 1 \) [11]. The reflection coefficients are scaled by a factor \( 0 < \alpha \leq 1 \). A value of \( \alpha \) close to 1 indicates that the reflected signal amount is more, and less energy is harvested for driving the circuitry. When \( \alpha \) is closer to 0, it means that the reflected signal is weak and will be effective only for very short-range communication. Hence, \( \alpha \) value has to be chosen judiciously such that enough power can be harvested for driving the circuitry and the reflected signal strength is good enough to reach the reader. In other words, \( \alpha \) is a backscattering parameter which determines the amount of signal power to be backscattered.

A 4-QAM backscattering tag can be physically realized with four lumped impedances that connect to an antenna port by Analog Devices ADG904 SP4T CMOS FET switch controlled by a TI MSP430F2011 microcontroller [11]. The microcontroller can be programmed to stream 128 bit strings of pseudo-random data to modulate the 4 switch states at 200k symbols/s, corresponding to a data rate of 400kbps. This microcontroller requires 3-6 mW of power for generating pseudo-random data. Hence, it can be powered by a CR2032 3V Lithium coin cell battery. The impedance modulation consumes only 115 nW of power. Similarly, a higher order 16-QAM backscattering tag can be practically realized by using a 16-to-1 multiplexer to switch the impedance between 16 modulation states. The modulation stage consumes 1.49 mW power in addition to the microcontroller and Lithium cell battery. Thus, the power consumption of our proposed system will be very low as compared to the power requirements of a traditional IoT device, which is around 50.11 mW [12].

B. Channel Modeling

In our system, channels correspond to three links: WiFi-to-tag, tag-to-RIS, and RIS-to-reader. Considering that the channel coefficients of the three links follow Nakagami distribution and are i.i.d., the overall channel model will represent a cascaded Nakagami distribution having 3 components. The
probability density function (PDF) of such a cascaded Nakagami model, where \( Y = \prod_{i=1}^{N} H_i \), can be written as [13]:

\[
 f_Y(y) = \frac{2}{y^{N} \prod_{i=1}^{N} \Gamma(m_i)} \left( m_1, m_2, \ldots, m_N \right) \left( y^{N} \prod_{i=1}^{N} \frac{m_i}{\epsilon_i} \right),
\]

where \( G_{m,n}^0(z) \) is Meijer-G function [14]. Random variable \( H_i \geq 0 \) follows a Nakagami distribution, whose PDF can be evaluated as shown in Eq. (1) of [13]. \( \Gamma(\cdot) \) is the Gamma function, \( m_i = \frac{(E[H_i])^2}{V_{avr}[H_i]} \), and \( \Omega_i = E[H_i^2] \), where \( E[\cdot] \) is the expectation operator and \( V_{avr}[\cdot] \) is the variance operator. The \( n^{th} \) order moment of \( Y \) can be evaluated as [13]:

\[
 E[Y^n] = \prod_{i=1}^{N} \Gamma(m_i + n/2) \left( \frac{\Omega_i}{m_i} \right)^{n/2},
\]

where \( n = 1 \) is used to calculate the mean. To get the variance, \( E[Y^2] \) is first calculated by considering \( n = 2 \). Then, variance is obtained as \( \sigma^2 = E[Y^2] - (E[Y])^2 \).

### III. Performance Analysis

#### A. Error Rate Analysis of Spatial Modulation

The reader having a single antenna receives the signal from the tag which employs SM. The resultant signal is given by:

\[
 y_{e2e}^{SM} = \alpha \sqrt{P_R} \left( \sum_{k=1}^{K} h_{i,k} v_i g_i \right) x + n_w, \tag{3}
\]

where \( P_R \) is the average transmit power of the WiFi router, and where \( h_{i,k} = d_1^{-u/2} \beta_i k e^{-j\psi_i} \) and \( g_i = d_2^{-u/2} \delta_i e^{-j\phi_i} \) are the channel gains of the tag-RIS and RIS-reader links, respectively. Since we apply SM or SSK to activate a single antenna, it is equivalent to selecting a particular \( k \)th column of the tag-RIS channel matrix. In (3), \( x \) is the modulated data symbol transmitted by the tag, which is inserted into the \( k \)th position and all entries are zero (i.e., data is transmitted through a single activated antenna). Also, \( d_1 \) and \( d_2 \) are the distances of the tag-RIS and RIS-reader links, respectively, and \( u \) is the path-loss exponent. Further, \( \beta_i, k \) and \( \delta_i \) denote the amplitudes following Nakagami distribution, whereas \( \psi_i \) and \( \phi_i \) are the phases. Besides, \( b_k \) denotes the \( k \)th column of the WiFi-tag channel matrix, whose elements are Nakagami distributed. The additive AWGN is denoted by \( n_w \). The \( i \)th element of the RIS induces a reflection coefficient given by \( \psi_i = \zeta_i(\Theta_i) e^{j\Theta_i} \). For ideal phase shifts, \( \zeta_i(\Theta_i) = 1 \) for any element. To achieve the maximum instantaneous signal-to-noise ratio (SNR), we set \( \Theta_i = \psi_i + \phi_i \) which eliminates the channel phases. Thus, assuming RIS as a perfect reflector, the received signal at the reader for SM can be modified as:

\[
 y_{e2e}^{SM} = \alpha \sqrt{P_R} \left( d_1 d_2 \right)^{-u/2} \sum_{k=1}^{K} \beta_i k \delta_i \left[ h_{i,k} v_i g_i \right] x + n_w, \tag{4}
\]

where \( \Xi = b_k \sum_{i=1}^{K} \beta_i k \delta_i \). The product of three Nakagami fading components gives rise to a cascaded Nakagami model whose mean and variance can be calculated as shown in subsection II.B, namely, mean \( \mu = \frac{\Gamma(m_i + 1/2)^2}{\Gamma(m_i)} \left( \frac{\Omega_i}{m_i} \right)^{3/2} \) and variance \( \sigma^2 = \frac{(\Gamma(m_i + 1)^2)}{\Gamma(m_i)} \left( \frac{\Omega_i}{m_i} \right)^{3} - \mu^2 \). Now, if we consider large samples of the channel coefficients having known mean and variance, then the distribution of the sample mean will be approximately normally distributed irrespective of the components distribution. This central limit theorem (CLT) is valid for our case as \( K >> 1 \). Thus, we can write \( \Xi \sim \mathcal{N}(\mu, K \sigma^2) \), where \( \mathcal{N}(\mu, \sigma^2) \) indicates the Gaussian distribution [3]. By applying maximum likelihood (ML) detection and assuming channel state information (CSI) is available, the antenna index \( m \) and the data symbol \( x \) can be jointly decoded as: \( (\hat{m}, \hat{x}) = \arg \min_{(m,x)} \left| y_{e2e}^{SM} - \theta \left( b_m \sum_{i=1}^{K} \beta_i m \delta_i \right) x \right|^2 \), \( \tag{5} \)

where \( \hat{m} \) and \( \hat{x} \) are the estimates of antenna index and modulated symbol, respectively, and \( \theta = \alpha \sqrt{P_R} \left( d_1 d_2 \right)^{-u/2} \). The pairwise error probability (PEP) for SM with \( N_i \) and \( N_t \) antennas at the reader and tag, respectively, is given by:

\[
 P(m, x \rightarrow \hat{m}, \hat{x}) = Q \left( \frac{1}{N_t} \sqrt{\sum_{i=1}^{N_t} \sum_{k=1}^{N_i} \left| G_{i,k} x - G_{i,k} \hat{x} \right|^2 / 2N_0} \right), \tag{6}
\]

where \( Q(x) = \frac{1}{\sqrt{2\pi}} \int_{x}^{\infty} \exp \left( -u^2 / 2 \right) du \) and \( N_0 \) is the AWGN power. Considering \( N_t = 1 \), Eq. (6) can be simplified as:

\[
 P(m, x \rightarrow \hat{m}, \hat{x}) = Q \left( \sqrt{\sum_{k=1}^{N_i} \left| G_{k} x - G_{k} \hat{x} \right|^2 / 2N_0} \right), \tag{7}
\]

where \( G_k = b_k \sum_{i=1}^{K} \beta_i k \delta_i \). Now, let \( \gamma_{eff} = \sum_{k=1}^{N_i} \left| G_{k} x - G_{k} \hat{x} \right|^2 \). Then, the PEP averaged over channel fading [9] can be calculated using

\[
 P(m, x \rightarrow \hat{m}, \hat{x}) = \int_{0}^{\infty} Q \left( \sqrt{\frac{\theta^2 \gamma_{eff}}{2N_0}} \right) f_{\gamma_{eff}}(\gamma_{eff}) d\gamma_{eff} \tag{8}
\]

where \( f_{\gamma_{eff}}(\gamma_{eff}) \) is the channel fading distribution. The PEP evaluation involves calculation of the moment generating function (MGF) of \( \gamma_{eff} \) in 2 stages depending on correct or erroneous detection of the antenna index.

1) Case I \( (m \neq \hat{m}) \): Let us assume \( \gamma_{eff} = \gamma_{eff_1} + \gamma_{eff_2} + \gamma_{eff_3} \), where \( \gamma_{eff_1}, \gamma_{eff_2}, \gamma_{eff_3} \) signify \( k = m, k = \hat{m}, \) and \( k \neq m, k \neq \hat{m} \) scenarios, respectively, evaluated as:

\[
 \gamma_{eff_1} = \left| \bar{b}_m \sum_{i=1}^{K} \beta_i m \delta_i \right| x - \left| \bar{b}_m \sum_{i=1}^{K} h_{i,m} g_i e^{j\Theta_i,m} \right| \hat{x} \right|^2 \tag{9}
\]

\( \text{CSI can be acquired with affordable overhead, e.g., by using matrix calibration based channel estimation [15].} \)
\[ \gamma_{\text{eff}} = \left| \mathbf{G}_m x - \hat{\mathbf{G}}_m \hat{x} \right|^2 \]
\[ \gamma_{\text{eff}} = \left( b_m \sum_{i=1}^K h_{i,m} g_i e^{j\theta_{i,m}} \right)^2 - \left( b_m \sum_{i=1}^K \beta_i m \delta_i \right) \hat{x}^2 \]
\[ \gamma_{\text{eff}} = \left[ \mathbf{G}_m x - \hat{\mathbf{G}}_m \hat{x} \right]^2 \]
\[ \gamma_{\text{eff}} = \sum_{k=1(k \neq m,k \neq \hat{m})}^{N_t} \left| \mathbf{G}_k x - \hat{\mathbf{G}}_k \hat{x} \right|^2. \quad (9) \]

Now, breaking down the channel coefficients, and after rearranging the terms, we can write:
\[ \gamma_{\text{eff}} = \left| \gamma_1 \right|^2 = (\gamma_1)^2 + (\gamma_2)^2 \]
\[ \gamma_{\text{eff}} = \left| \gamma_2 \right|^2 = (\gamma_2)^2 + (\gamma_3)^2 \]
\[ \gamma_{\text{eff}} = \sum_{k=1(k \neq m,k \neq \hat{m})}^{N_t} \left| \mathbf{G}_k \right|^2. \quad (10) \]

where \( \Theta_i = \Theta_{i,m} - \Theta_{i,\hat{m}} \). The phase angles \( \Theta_i \) are independently and uniformly distributed in \((0, 2\pi)\). Hence, the difference of these phases has a triangular PDF:
\[ f_{\Theta}(x) = \begin{cases} \frac{1}{2\pi} (1 + \frac{x}{2\pi}), & -2\pi < x < 0 \\ \frac{1}{2\pi} (1 - \frac{x}{2\pi}), & 0 < x < 2\pi. \end{cases} \]

From the CLT, we can state that \( \gamma_1 \) and \( \gamma_2 \) are Gaussian distributed for large \( K \) values. We consider \( \gamma_{\text{eff}} = \mathbf{g}^T \mathbf{g} \) for \( \mathbf{g} = [(\gamma_1)_{\mathbf{R}} (\gamma_2)_{\mathbf{R}} (\gamma_2)_{\mathbf{R}}]$$. In which \( \mathbf{I} \) is an identity matrix of size \( 4 \times 4 \) and \( [\mathbf{g}]^T \) defines the transpose of a matrix. The mean vector of \( \mathbf{g} \) is \( \mathbf{M} = [\mu_{\mathbf{R}} \mu_{\mathbf{R}} - \mu_{\mathbf{R}} \mu_{\mathbf{R}}]^T \), while the covariance matrix \( \mathbf{C} \) of \( \mathbf{g} \) is
\[ \mathbf{C} = \begin{bmatrix} \sigma_1^2 & \sigma_1 \sigma_2 & \sigma_1 \sigma_3 & \sigma_1 \sigma_4 \\
\sigma_2 \sigma_1 & \sigma_2^2 & \sigma_2 \sigma_3 & \sigma_2 \sigma_4 \\
\sigma_3 \sigma_1 & \sigma_3 \sigma_2 & \sigma_3^2 & \sigma_3 \sigma_4 \\
\sigma_4 \sigma_1 & \sigma_4 \sigma_2 & \sigma_4 \sigma_3 & \sigma_4^2 \end{bmatrix}. \quad (12) \]
\[ \mathbf{M} \) and \( \mathbf{C} \) have dimensions of \( 4 \times 1 \) and \( 4 \times 4 \), respectively.

Adopting the approach in [9], the elements of \( \mathbf{C} \) are calculated as:
\[ \sigma_1^2 = K(\hat{x}^2 + 0.5|\hat{x}|^2), \quad \sigma_2^2 = K(\hat{x}^2 + 0.5|\hat{x}|^2), \quad \sigma_3^2 = K(\hat{x}^2 + 0.5|\hat{x}|^2), \quad \sigma_4^2 = K(\hat{x}^2 + 0.5|\hat{x}|^2) \]

The cross-correlation coefficients are calculated as \( c_{ij} = (M(i)\sigma_j^2 + M(j)\sigma_i^2)/(\sigma_i^2 + \sigma_j^2) - M(i)M(j) \). Thus, the MGF of \( D = \gamma_{\text{eff}} \) as evaluated as:
\[ M_D(s) = \left( \det \left( \mathbf{I} - 2s \mathbf{C} \right) \right)^{-\frac{1}{2}} \times \exp \left( -\frac{1}{2} \mathbf{M}^T \left[ \mathbf{I} - (\mathbf{I} - 2s \mathbf{C})^{-1} \right] \mathbf{C}^{-1} \mathbf{M} \right). \]
\[ \gamma_{\text{eff}} = \sum_{k=1(k \neq m,k \neq \hat{m})}^{N_t} \left| \mathbf{G}_k \right|^2. \quad (13) \]

Now from Eq. (9), we can write:
\[ \gamma_{\text{eff}} = \sum_{k=1(k \neq m,k \neq \hat{m})}^{N_t} \left| \mathbf{G}_k \right|^2 \]

According to the CLT, for large \( K \) values, \( \gamma_{\text{eff}} \) can be written as the sum of \( N_t - 2 \) independent central Chi-square random variables with 2 degrees of freedom, and its MGF can be expressed as
\[ M_{\gamma_{\text{eff}}}(s) = \left( \frac{1}{1 - sK(|\hat{x}|^2 + |\hat{x}|^2)} \right)^{N_t - 2}. \quad (15) \]

By multiplying Eq. (13) and (15), we obtain the overall MGF of \( \gamma_{\text{eff}} \) and can thereafter compute the PEP shown in Eq. (8). The end results are not shown here due to lack of space, and because the final calculation steps are straightforward.

2) \( \text{Case II} (m = \hat{m}) \)

Now, we compute the PEP for the scenario when the antenna index has been correctly detected. Considering \( \mathbf{G}_k = \mathbf{G}_k \), we can write:
\[ \gamma_{\text{eff}} = \sum_{k=1(k \neq m,k \neq \hat{m})}^{N_t} \left| \mathbf{G}_k \right|^2 = |x - \hat{x}|^2 \left( \mathbf{G}_m + \sum_{k=1(k \neq m)}^{N_t} |\mathbf{G}_k|^2 \right). \quad (16) \]

It is pertinent to note that \( \mathbf{G}_m \sim \mathcal{N}(K\mu, K\sigma^2) \) and \( \mathbf{G}_k \sim \mathcal{N}(0, K) \) for \( k \neq m \). \( \mathbf{G}_m \) follows non central Chi-square distribution with one degree of freedom, while \( \mathbf{G}_k \) follows central Chi-square distribution with two degrees of freedom. Therefore, the overall MGF can be calculated as:
\[ M_{\gamma_{\text{eff}}}(s) = \left( \frac{1}{1 - 2sK\sigma^2 |x - \hat{x}|^2} \right)^\frac{1}{2} \exp \left( \frac{sK^2 |x - \hat{x}|^2 \mu^2}{2sK |x - \hat{x}|^2} \right) \]
\[ \times \left( \frac{1}{1 - sK |x - \hat{x}|^2} \right)^{N_t - 1}. \quad (17) \]
Substituting the value from Eq. (17) in Eq. (8), we get the final PEP. The upper-bound BER is calculated as:
\[ P_b^{\text{SM}} = \frac{1}{MN_t} \sum_m \sum_{\hat{m}} \sum_x \sum_{\hat{x}} \tilde{P}(m, x \rightarrow \hat{m}, \hat{x}) e(m, x \rightarrow \hat{m}, \hat{x}), \quad (18) \]

where \( e(m, x \rightarrow \hat{m}, \hat{x}) \) denotes the number of erroneous bits encountered while calculating the pairwise vectors.

B. Error Rate Analysis of Space Shift Keying

For the system operation with SSK, the received signal at the reader is given by:
\[ y_{e}\text{SK} = \alpha \sqrt{P_R} \left[ \sum_{i=1}^K b_k h_{i,k} v_i g_i \right] + n_w. \quad (19) \]

The received signal can be modified in a similar manner as shown in Eq. (4) and rewritten as:
\[ y_{e}\text{SK} = \alpha \sqrt{P_R(d_1 d_2)}^{-u/2} \mathbf{\Xi} + n_w. \quad (20) \]

Now, the instantaneous SNR at the reader is given by:
\[ \gamma_{e}\text{SK} = \frac{\alpha^2 P_R \mathbf{\Xi}}{N_0(d_1 d_2) u}. \quad (21) \]

By applying ML detection, the antenna index is decoded as:
\[ \hat{m} = \arg \min_m \left| y_{e}\text{SK} - \varphi \left[ b_m \sum_{i=1}^K \beta_i m \delta_i \right] \right|^2. \quad (22) \]
For SSK, the PEP conditioned on the channel coefficients is:

\[ P(m \rightarrow \hat{m}) = Q \left( \sqrt{\frac{\alpha^2 \sum_{k=1}^{N_t} \left| G_k - \hat{G}_k \right|^2}{2N_0}} \right). \]  

(23)

The unconditional PEP evaluation, \( \hat{P}(m \rightarrow \hat{m}) \), and the MGF computation can be done similarly to the case with SM. Here for SSK, the symbol error case does not arise. Hence, only Case I is used for MGF calculation and the final BER upper-bound can be simplified as \( P_{\text{upper bound}}^{\text{SSK}} \leq \frac{N_t}{2} P(m \rightarrow \hat{m}) \).

C. Outage Probability Analysis

The outage probability for both operation scenarios, i.e., with SSK or SM, remains the same for a threshold data rate of \( R \) bps, and is given by

\[ P_{\text{out}} = 1 - Q \left( \frac{\sqrt{\lambda}}{\kappa}, \frac{\sqrt{\gamma_{\text{th}}}}{\kappa} \right), \]  

(24)

where \( Q_c(a, b) \) is the Marcum Q-function [14], [16], \( \lambda = \gamma e^{2\mu^2} \), \( \kappa = \sqrt{\gamma e^{2\sigma^2}} \), and \( \gamma_{\text{th}} = 2^{(R - \log_2(N_t))} - 1 \).

IV. Numerical Results and Comparisons

In the simulation experiments, the parameters considered are \( \alpha = 0.3, M = 4, a_1 = 0.2 \) Km, \( a_2 = 0.5 \) Km, \( u = 2 \), \( m_1 = m_2 = m_3 = 2 \), \( \Omega_1 = \Omega_2 = \Omega_3 = 1 \), \( N_t = 2 \), and \( K = 32 \). These parameters are valid for all the results unless mentioned explicitly. Such a setting is useful for indoor applications where the tag and WiFi router located nearby in a room can transmit to RIS (designed on the room walls) situated slightly farther, and the reader is located much farther in a different room. All the analytical upper-bound results are corroborated with Monte Carlo simulation results.

Fig. 2: BER for different numbers of the RIS elements.

The BER results of RIS-assisted ABSc communications are compared in Fig. 2 for the cases with SM or SSK. The number of RIS elements (\( K \)) is varied as 16, 32 and 64. As observed, an increase in the number of RIS elements improves the BER performance for both SSK and SM. SSK performs slightly better than SM due to the fact that symbol modulation error is not present in SSK. However, SSK has a lower SE than SM. Hence, a trade-off between SE and BER performance is required. It is pertinent to note that SSK and SM for any number of RIS elements also perform better than the scenario where only SM and ABSc techniques are applied without RIS. Such a scenario considers \( \alpha = 0.3 \) and \( u = 2 \), but the distance between the tag and the reader is only 20 m. Our proposed SM and SSK methods aided by RIS are capable of improving the BER performance at a much larger distance of separation which is possible due to the scaling property of RIS. It is also observed that our proposed methods perform better than conventional \( M \)-QAM based data transfer utilizing ABSc and RIS (with \( K = 32 \)) without involving SM or SSK (cf. plot marked as ABSc, RIS, no SM). In particular, SSK achieves a SNR gain of minimum 2 dB over SM for any number of RIS elements. For any SNR value of more than -20 dB, there is a SNR gain of minimum 6 dB between any number of RIS elements for either SSK or SM. SM and SSK in the presence of RIS are capable of attaining a minimum SNR gain of 10 dB over the system without RIS.

Fig. 3: Effect of the modulation and tag antennas on BER.

The influence of modulation scheme and the number of tag antennas are investigated in Fig. 3. Higher modulation schemes lead to degradation of SM performance for a fixed number of RIS elements. This is due to the fact that the SE increases for higher modulation levels, and higher SE values are difficult to be achieved. With an increment in \( N_t \), the plots have to be regenerated for different modulation levels, which improves the performance of both SM and SSK by a huge margin. For SSK, \( N_t = 8 \) achieves a SNR gain of minimum 4 dB over \( N_t = 4 \), and a minimum of 8 dB over \( N_t = 2 \) for SNRs higher than -20 dB. Similarly, for SM, \( N_t = 8 \) achieves a SNR gain of minimum 4 dB over \( N_t = 4 \) and a minimum of 10 dB over \( N_t = 2 \) for high SNRs of more than -20 dB. However, in case of lower SNR values, the performance with larger \( N_t \) is inferior for both SSK and SM. But the presence of RIS results in higher received SNR in the high-SNR regime, thereby nullifying the presence of more tag antennas at higher SNR values, which was not possible in case of conventional SM without RIS.

The effects of different parameters on the system perfor-
performance are explored in Fig. 4. Here, Case 1 corresponds to SM with \( K = 32, M = 4, \alpha = 0.3, u = 2, d_1 = 0.2 \text{ Km}, \) \( d_2 = 0.5 \text{ Km}; \) Case 2 corresponds to SM with \( K = 32, M = 4, \alpha = 0.3, u = 2, d_1 = d_2 = 0.5 \text{ Km}; \) Case 3 represents SM with \( K = 32, M = 4, \alpha = 0.5, u = 2, d_1 = 0.5 \text{ Km}; \) and Case 4 denotes SM with \( K = 32, M = 4, \alpha = 0.3, u = 4, d_1 = d_2 = 0.5 \text{ Km}. \) Meanwhile, Case 5 corresponds to SSK with \( K = 32, \alpha = 0.3, u = 2, d_1 = 0.2 \) Km, \( d_2 = 0.5 \text{ Km}; \) Case 6 denotes SSK with \( K = 32, \alpha = 0.3, u = 2, d_1 = d_2 = 0.5 \text{ Km}; \) Case 7 indicates SSK with \( K = 32, \alpha = 0.5, u = 2, d_1 = d_2 = 0.5 \text{ Km}; \) and Case 8 represents SSK with \( K = 32, \alpha = 0.3, u = 4, d_1 = d_2 = 0.5 \text{ Km}. \) It is evident from the figure that shorter distances and larger \( \alpha \) are beneficial for SSK and SM performance in the presence of RIS. Larger \( \alpha \) means that more power gets reflected from the backscatter link, hence the signal power at the RIS is more. Larger distances mean inferior received signal quality, leading to poor BER performance. For all the plots, the analytical upper bounds are tight for high SNR values.

The outage probability results are plotted for \( K = 16 \) and 32 RIS elements in Fig. 5. It is seen that higher \( K \) and lower data rate values \( (R) \) yield better performance. Higher data rates are difficult to achieve, leading to more outage events. The outage probability tends to reach very low values for larger SNRs because the RIS results in higher received SNR.

V. CONCLUSION

A novel spectrally and energy efficient SM and SSK-based RIS system implementing ABSc technology has been proposed in this paper for effective long-distance communications between 700 m-1000 m. The performance evaluation demonstrated that SM and SSK in the presence of RIS are capable of achieving a SNR gain of minimum 10 dB over the system without RIS for higher SNR values. The derived analytical upper bounds were shown to be tight for high SNR values. Future work includes the implementation of multiple RIS, and multiple antennas at the reader. Advanced SM schemes can also be applied both at the tag and the RIS(s) to select multiple antennas or RIS reflecting elements.

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