Intelligent Reflecting Surface-Aided Backscatter Communications

Xiaolun Jia*, Jun Zhao†, Xiangyun Zhou* and Dusit Niyato‡

*Research School of Electrical, Energy and Materials Engineering, The Australian National University, Canberra, Australia
†School of Computer Science and Engineering, Nanyang Technological University, Singapore
‡Email: xiaolun.jia@anu.edu.au, junzhao@ntu.edu.sg, xiangyun.zhou@anu.edu.au, dniyato@ntu.edu.sg

Abstract—We introduce a novel system setup where a backscatter device operates in the presence of an intelligent reflecting surface (IRS). In particular, we study the bistatic backscatter communication (BackCom) system assisted by an IRS. The phase shifts at the IRS are optimized jointly with the transmit beamforming vector of the carrier emitter, to minimize the transmit power consumption at the carrier emitter, whilst guaranteeing a required BackCom performance. The unique channel characteristics arising from multiple reflections at the IRS render the optimization problem highly non-convex. Therefore, we utilize the minorization-maximization (MM) algorithm and the semidefinite relaxation (SDR) technique, and present an approximate solution for the optimal IRS phase shift design. We also extend our analytical results to the monostatic BackCom system. Numerical results indicate that the introduction of the IRS brings about considerable reductions in transmit power, even with moderate IRS sizes, which can be translated to range increases over the non-IRS-assisted BackCom system.

I. INTRODUCTION

Backscatter communication, or BackCom, has received increasing research interest in recent times, as a potential solution to address the energy efficiency and sustainability of sensor networks under the Internet of Things (IoT). Conventional applications of BackCom include radiofrequency identification (RFID) systems, where RFID tags transmit small data packets to a reader by performing modulation on an existing signal. The concept of backscatter by reflection has since emerged as a key technology for industrial IoT and pervasive wide-area networking: as a result, the bistatic [1] and ambient [2] architectures have since been proposed, to improve the range of BackCom and its compatibility with radiofrequency (RF) signals that are already modulated.

Despite the large amount of literature on improving BackCom system performance in terms of reliability and throughput, the short range of BackCom is still a prominent barrier preventing its widespread deployment. Specifically, the signal power received from BackCom devices, or tags, in monostatic systems, scales inversely with the fourth power of the tag-reader distance. Bistatic systems require dedicated carrier emitters (CEs) placed close to tags in low-interference environments to achieve longer range; and ambient systems suffer from direct-link interference, which incurs high complexity to mitigate, thereby limiting its range to a few meters.

Intelligent reflecting surfaces (IRS) have been recently proposed as a means to modify the wireless propagation medium. An IRS consists of a number of reflecting elements, each made from metamaterials, which interact directly with impinging signals to alter their characteristics such as amplitude and phase. In addition to being almost passive, coordinating a large number of reflectors at an IRS in jointly designing their phase shifts allows reflected signals to be received constructively (or destructively) at a receiving node [3]. This allows for favorable SNR scaling at the receiver in the case of constructive reflection, where the SNR is shown to scale with the square of the IRS surface area [4].

As a result, many works have examined the performance improvements of introducing IRS to a range of communication systems. Work in [5], [6] studied fundamental metrics of IRS-assisted systems, such as error performance and capacity; while detailed analysis of propagation and path loss in IRS-reflected links were provided in [4], [7]. The joint optimization of IRS phase shifts and transmit beamforming vectors were examined in works such as [8], [9], among many others. More recently, research attention has shifted towards facilitating joint energy and information transfer using IRS through works such as [10], [11]. However, it should be noted that the use of IRS to support passive communication, specifically BackCom, has received relatively little attention in the literature.

In this paper, we study a system where a BackCom device operates in the presence of a nearby IRS. In light of the rapid uptake in IRS research and its expected widespread deployment, it is necessary to consider the performance of existing communication systems, with BackCom being an example, where an IRS is likely to be nearby. To the best of our knowledge, the only other work jointly considering IRS and detached BackCom devices is [12]. However, [12] studies the error performance of a non-conventional monostatic system where no direct link exists between the reader and tag; whereas no works have considered the standard monostatic or bistatic BackCom architectures assisted by IRS. The main contributions of this paper are as follows:

- We introduce an IRS-aided BackCom system where the backscatter communication from the tag to the reader is assisted by the IRS. This new BackCom system has very unique channel characteristics, where signals may be reflected multiple times by the IRS.
- Specifically considering a bistatic BackCom system where a tag’s backscatter communication to a reader is powered by a multi-antenna CE, the presence of the IRS reflects the signals from both the CE and the tag. This
presents a highly non-trivial design problem on the IRS phase shifts. We jointly optimize the IRS phase shifts and the transmit beamforming at the CE in order to minimize the required transmit power consumption at the CE.

- We further extend our analysis to a monostatic BackCom system, and obtain the optimal phase shifts for the IRS.
- Numerical results reveal notable reductions in the required transmit power compared to monostatic and bistatic BackCom systems without IRS, which can be translated to improvements in the link budget and range.

**Notations:** $j = \sqrt{-1}$ denotes the complex unit, and $\mathbb{C}$ denotes the set of complex numbers. $|\cdot|$ and $\text{Re}\{\cdot\}$ denote the magnitude and the real part of a complex number, respectively. $\mathcal{CN}(\mu, \sigma^2)$ represents a complex Gaussian distribution with mean $\mu$ and variance $\sigma^2$. Vector and matrix quantities are denoted using lowercase and uppercase boldface letters, respectively, as in $a$ and $A$. $I$ denotes the identity matrix of variable size, depending on the context. $\|a\|$ denotes the Euclidean norm of a vector; and $\text{Tr}(A)$ and $A^H$ denote the trace and the Hermitian transpose of $A$, respectively.

II. SYSTEM AND SIGNAL MODEL

A. System Setup

We consider an IRS-aided bistatic BackCom system with an $L$-antenna CE, a single-antenna tag, a single-antenna reader, and an IRS with $N$ reflecting elements. Hereafter, we denote the CE, tag, IRS and reader by $C$, $T$, $I$ and $R$, respectively. A system diagram is shown in Fig. 1. Despite the simplicity of the system, the design complexity arises from the fact that the IRS must balance the signal reflections from both the CE-to-tag and tag-to-reader links. The associated phase shift design problem is highly nonconvex, and is explored in Section III.

The CE transmits a continuous-wave signal with maximum power $P$ and beamforming vector $w \in \mathbb{C}^{L \times 1}$ to power the tag’s communication, such that $\|w\|^2 \leq P$.

The tag has $J$ load impedances connected to its antenna. We assume that the tag has an off-state where the load and antenna impedances are perfectly matched, resulting in no reflection and no data transmission. We consider a generalized tag configuration where the tag could be either passive or semipassive, with circuit power consumption $\xi$. Where $\xi = 0$, the tag is powered by an on-board battery; otherwise, a portion of energy from the incoming signal is used to power the tag.

Each reflecting element of the IRS is modeled as a diffuse reflector $[4]$, $[7]$. As diffuse reflections incur significant path loss, we ignore signals which undergo two or more reflections at the tag. However, since the IRS is able to tune all of its reflectors to focus the signal at a receiver, this assumption does not apply to signals reflected exactly two times by the IRS. That is, the C-I-T-I-R link is still considered.

We consider the ideal assumption that the reader has perfect knowledge of the channel state information (CSI) of all channels. The perfect CSI assumption allows us to characterize the upper bound on the system performance. Note that the evaluation of any channel estimation methods for an IRS-aided BackCom system is outside the scope of this work.

B. Signal Model

The tag modulates its data onto an incident signal by switching between its impedances $[1]$, $[2]$. Its baseband signal is denoted by $b(t)$, which is the tag’s time-varying reflection coefficient. $b(t)$ is the reflection coefficient takes on values $b(t) \in \{b_1, \ldots, b_J\}$, where each value corresponds to an impedance; moreover, $|b_i| = |b_j|$, $\forall i, j$, and $|b_i| \leq 1$, $\forall i$. The power splitting coefficient at the tag is denoted by $\alpha \in [0, 1]$, where $\alpha$ denotes the proportion of the signal to be reflected, with $1-\alpha$ proportion of the signal used to power the circuit.

The continuous-wave signal transmitted by the CE is given by $s(t)$. Denote the channels from the CE to tag, CE to IRS, CE to reader, IRS to tag, IRS to reader and tag to reader by $h_{CT} \in \mathbb{C}^{1 \times L}$, $h_{CI} \in \mathbb{C}^{N \times L}$, $h_{CR} \in \mathbb{C}^{1 \times L}$, $h_{IT} \in \mathbb{C}^{1 \times N}$, $h_{IR} \in \mathbb{C}^{1 \times N}$ and $h_{TI} \in \mathbb{C}^{1 \times 1}$, respectively. Each IRS element $n \in \{1, \ldots, N\}$ reflects the sum of all incident signals with a phase shift, denoted by $\theta_n$. We assume the amplitude gain of each IRS reflector to be unity. Let the vector containing the phase shifts of all reflectors be $\theta = [\theta_1, \ldots, \theta_N]$, where $\theta_n \in [0, 2\pi)$ $\forall n$. The matrix of IRS phase shifts can then be written as $\Theta = \text{diag}(e^{j\theta_1}, \ldots, e^{j\theta_N})$.

Linear transmit precoding is assumed at the CE, where a single beamforming vector $w$ is used. The transmitted signal can then be written as $x_C = ws(t)$. The signal received at the tag consists of the direct C-T link and the reflected C-I-T link, and can be written as

$$y_T = (h_{CT}^H \Theta h_{CI} + h_{CT}) w s(t). \quad (1)$$

There is no noise term at the tag, as no signal processing is performed. The part of the signal reflected by the tag is

$$x_{T,I} = \sqrt{\alpha} b(t) y_T. \quad (2)$$

The remainder of the signal is harvested to power the circuit, whose squared magnitude, denoted by $E_h$, can be given by

$$E_h = \eta (1 - \alpha)|y_T|^2; \quad (3)$$

$^4$Realistically, the tag baseband signal is of the form $b(t) = A - \Gamma(t)$, where $\Gamma(t)$ is the reflection coefficient, and $A$ is the tag’s antenna structural mode, and determines the default amount of signal reflection in the off-state. However, $A$ is a constant, and can hence be subtracted from the received signal in post-processing. Therefore, we do not consider it here.
where $\eta \in [0, 1]$ is the energy harvesting efficiency. As a result, the circuit constraint is given by

$$(1 - \alpha)\eta \left( |h_{TI}^H \Theta H_{CI} + h_{CT}| \right)^2 \geq \xi,$$

with $\xi$ being the circuit power consumption in Watts. The signal received at the reader consists of those arrived directly from the CE, backscattered from the tag, and reflected signal received at the reader consists of those arrived directly from the CE, backscattered from the tag, and reflected with $
_T \in \mathbb{C}$ divided by $\sigma_R^2$, using the received signal power when the tag is in its off-state as reference.

### III. Transmit Power Minimization Problem

In this paper, we study the transmit power minimization problem at the CE, subject to an SNR constraint for the tag’s information transmission to the reader. The transmit power minimization problem is appealing for a BackCom system, as it allows not only the determination of optimal IRS and BackCom device parameters, but also the possibility of translating the power saving from the optimal solution to an increase in the power budget and hence range.

To solve the problem, we jointly optimize the transmit beamforming vector at the CE, the phase shift coefficients at the IRS, and the power splitting coefficient at the tag. We begin by presenting the problem for a bistatic BackCom system, and extend it to the monostatic architecture in the sequel. The problem can be written as (Problem $P$):

$$\min_{P, \alpha, \Theta} P \quad \text{s.t.} \quad (6a)$$

$$\min_{P, \Theta, \alpha} \|w\|^2 \quad \text{s.t.} \quad (6a)$$

$$0 \leq \alpha \leq 1,$$

$$0 \leq \theta_n \leq 2\pi, \forall n \in \{1, \ldots, N\},$$

where $(6a)$ is the tag’s SNR constraint, $(6b)$ is the tag’s circuit power constraint, $(6c)$ is the splitting coefficient constraint, and $(6e)$ is the phase shift constraint for each IRS reflector.

#### A. Minimum Transmit Power Solution

We first simplify Problem $P$, and then obtain the expression for the minimum transmit power. For conciseness, the following substitutions are made in equations hereafter:

$$H_1(\Theta) \triangleq h_{TI}^H \Theta H_{CI} + h_{CT},$$

$$H_2(\Theta) \triangleq h_{TI}^H \Theta h_{TI} + h_{TR}.$$  

$(7)$ and $(8)$ correspond to the combined channel gains for the CE-to-tag and tag-to-reader links, respectively, which include the reflected signal paths from the IRS.

We begin by noting that when there is one user, optimal beamforming can be achieved using maximum ratio transmission (MRT). Therefore, the optimal $w$ in Problem $P$ is simply

$$w^* = \sqrt{\frac{P}{\|H_2(\Theta)H_1(\Theta)\|^2}}.$$  

Substituting $(9)$ into Problem $P$, we can rewrite it as a power minimization problem (Problem $P_1$):

$$\min_{P, \alpha, \Theta} P \quad \text{s.t.} \quad (10a)$$

$P = \min_{P, \alpha, \Theta} \quad \text{s.t.} \quad (10b)$

$P(1 - \alpha)\eta \|H_1(\Theta)\|^2 \geq \xi,$$

$P(10c)$

$(6d)$

$(6e)$

By inspection, the minimum transmit power is the minimum value that would meet both the SNR and circuit power constraints with equality. From $(10b)$ and $(10c)$, an intermediate expression of the minimum transmit power $P^*$ is

$$P^* = \max \left\{ \frac{\xi}{(1 - \alpha)\eta \|H_1(\Theta)\|^2}, \left( \frac{\gamma_{th}\sigma_R^2}{\xi \|H_2(\Theta)H_1(\Theta)\|^2} \right)^{1/2}, \left( \frac{\eta \gamma_{th}\sigma_R^2}{\xi \|H_2(\Theta)H_1(\Theta)\|^2} \right)^{1/2} \right\}.$$  

From $(11)$, the optimal splitting coefficient $\alpha^*$ can be found.

Note that as $\alpha$ is increased from 0 to 1, the first term of $(11)$ is monotonically decreasing; while the second term of $(11)$ is monotonically increasing. Therefore, the optimal value of $\alpha^*$ that minimizes $P$ is found by equating the two terms:

$$\alpha^* = \frac{\eta \gamma_{th}\sigma_R^2}{\xi \|H_2(\Theta)H_1(\Theta)\|^2}.$$  

The minimum transmit power is then found by substituting $\alpha^*$ into either term in $(11)$:

$$P^* = \frac{\eta \gamma_{th}\sigma_R^2}{\xi \|H_2(\Theta)H_1(\Theta)\|^2}.$$  

$(12)$

$(13)$

To numerically quantify $P^*$, the optimal IRS phase shifts $\Theta$ need to be determined. Depending on the value of $\xi$, different methodologies are needed to minimize $(13)$. If $\xi = 0$, we can directly maximize the denominator over $\Theta$, as per the following problem:

$$\max_{\Theta} \quad \|H_2(\Theta)H_1(\Theta)\|^2,$$

$s.t. \quad (6e).$  

Otherwise, $\Theta$ appears in the numerator and denominator of $(13)$; and fractional programming (FP) techniques are needed.

#### B. IRS Phase Shift Design: No Circuit Power Constraint

In this subsection, we propose a solution to the IRS phase shifts where the tag is semipassive. Note that $H_2(\Theta)$ is a complex scalar. Therefore, Problem $(14)$ is similar to $8$.
As a result, denoting the original objective function in (15)
with one significant difference being an additional
$|H_2(\Theta)|^2$ term outside (Problem P2-nc):
\[
\max_\theta \ |H_2(\Theta)|^2 \left\| H_1^T \Theta H_{CI} + h_{CT} \right\|^2,
\]
s.t. \quad 0 \leq \theta_n \leq 2\pi, \forall n \in \{1, \ldots, N\}.
\tag{15} \]

First, \( |H_1^T \Theta H_{CI} + h_{CT}|^2 \) can be rewritten as follows:
\[
\left\| H_1(\Theta) \right\|^2 = v^H \Phi_{CIT} \Phi_{CIT}^H v + v^H \Phi_{CIT} h_{CT}^H + h_{CT} \Phi_{CIT}^H v + \left\| h_{CT} \right\|^2,
\]
where \( \Phi_{CIT} \triangleq \text{diag}(h_{CI}^T) H_{CI} \) and \( v \triangleq [e^{j\theta_1}, \ldots, e^{j\theta_N}]^H \), with \( |v_n|^2 = 1, \forall n \). It is evident that (16) is of quadratic form, and can therefore be rewritten in matrix form as
\[
\left\| H_1(\Theta) \right\|^2 = \bar{v}^H R \bar{v} + \left\| h_{CT} \right\|^2,
\]
with
\[
R = \begin{bmatrix} \Phi_{CIT} \Phi_{CIT}^H & \Phi_{CIT} h_{CT}^H \\ h_{CT} \Phi_{CIT}^H & 0 \end{bmatrix}, \quad \bar{v} = \begin{bmatrix} v \\ 1 \end{bmatrix}.
\tag{18} \]

With the additional \( |H_2(\Theta)|^2 \) term, it is clear that the problem under the BackCom system is considerably different to that in [8], which simply optimizes the objective function in [17]. The expanded form of \( |H_2(\Theta)|^2 \) can be given by
\[
\left\| H_2(\Theta) \right\|^2 = \bar{v}^H S \bar{v} + \left\| h_{TR} \right\|^2,
\]
with \( \Phi_{TIR} \triangleq \text{diag}(h_{TR}^T) h_{TR}^H \). This equation can also be rewritten in quadratic form as follows:
\[
\left\| H_2(\Theta) \right\|^2 = \bar{v}^H S \bar{v} + \left\| h_{TR} \right\|^2,
\tag{19} \]
with
\[
S = \begin{bmatrix} \Phi_{TIR} \Phi_{TIR}^H & \Phi_{TIR} h_{TR}^H \\ h_{TR} \Phi_{TIR}^H & 0 \end{bmatrix}.
\tag{20} \]

As a result, denoting the original objective function in \( \text{[15]} \) by \( F \), the product of (17) and (20) can be written as
\[
F(\bar{v}) = (\bar{v}^H R \bar{v} + c_1)(\bar{v}^H S \bar{v} + c_2)
= \bar{v}^H S \bar{v} \bar{v}^H R \bar{v} + c_1 \bar{v}^H S \bar{v} + c_2 \bar{v}^H R \bar{v} + c_1 c_2,
\tag{21} \]
with \( c_1 = \|h_{CT}\|^2 \) and \( c_2 = \|h_{TR}\|^2 \). (22) is a quartic polynomial in \( \bar{v} \). Normally, to optimize (17), we can let \( \bar{V} \triangleq \bar{v}^H \), and use the identity \( \bar{v}^H R \bar{v} = \text{Tr}(R \bar{v} \bar{v}^H) \) to recast (17) as a function of \( V \), which is rank-one. However, here we cannot invoke the trace identity, as the resulting trace is quadratic in \( V \) and is generally nonconvex. It has also been noted in [13] that it is NP-hard to optimize (minimize) polynomials of degree 4, meaning that a closed-form, optimal solution to (22) is generally not available.

To address this challenging issue, we resort to the semidefinite relaxation (SDR) technique nested within an MM algorithm, where we find a convex minorizing function to (22). In each iteration, we first obtain a relaxed solution to the objective function \( F(\bar{v}) \). Then, randomization is performed to ensure the solution adheres to the rank-one constraint. The process is repeated until convergence of the MM algorithm.

In a similar manner compared to [14] Lemma 12], we can construct a minorizer to a real-valued function \( f \) of complex variable \( x \) and bounded curvature by taking the second-order Taylor expansion with a negative squared error term:
\[
f(x) \geq f(x_0) + \text{Re} \left\{ \nabla f(x_0)^H (x - x_0) \right\} - \frac{L}{2} \| x - x_0 \|^2,
\tag{23} \]
where \( x_0 \in \mathbb{C}^N \) is any point, and \( L \) is the maximum curvature of \( f(x) \). Applying (23) to (22), we obtain
\[
F(\bar{v}) \geq \bar{v}_0^H S \bar{v}_0 \bar{v}_0^H R \bar{v}_0 + c_1 \bar{v}_0^H S \bar{v}_0 + c_2 \bar{v}_0^H R \bar{v}_0 + c_1 c_2
+ \bar{v}_0^H (\bar{v} - \bar{v}_0) + (\bar{v} - \bar{v}_0)^H T \bar{v}_0
\]
\[
- \frac{L}{2} (\bar{v}^H \bar{v} - \bar{v}^H \bar{v}_0 - \bar{v}_0^H \bar{v} + \| \bar{v}_0 \|^2)
\]
\[
= \frac{L}{2} (\bar{v}^H \bar{v} - \bar{v}^H \bar{v}_0 - \bar{v}_0^H \bar{v} + \| \bar{v}_0 \|^2)
\]
\[
+ \bar{v}_0^H T \bar{v} + \bar{v}^H T \bar{v}_0 + c
\]
\[
= \frac{L}{2} (\bar{v}^H I \bar{v} + \bar{v}^H \left( -\frac{2}{L} T \bar{v}_0 - I \bar{v}_0 \right)
\]
\[
+ \left( -\frac{2}{L} T \bar{v}_0 - I \bar{v}_0 \right)^H \bar{v} + c,
\tag{24} \]
where we define \( T \triangleq R \bar{v}_0 \bar{v}_0^H S + S \bar{v}_0 \bar{v}_0^H R + c_2 R + c_1 S \), which is a Hermitian matrix; \( c \) denotes the cumulative sum of all constant terms and terms involving only \( \bar{v}_0 \). The right-hand side of (24) is of quadratic form, and can be rewritten in matrix form as \( \bar{v}^H U \bar{v} \), where
\[
U = \begin{bmatrix} I & -\frac{2}{L} T \bar{v}_0 - I \bar{v}_0 \\ 0 & -\frac{2}{L} T \bar{v}_0 - I \bar{v}_0 \end{bmatrix}, \quad \bar{v} = \begin{bmatrix} \bar{v} \\ 1 \end{bmatrix}.
\tag{25} \]

Then, letting \( \tilde{V} = \bar{v} \bar{v}^H \), Problem P2-nc can now be rewritten as (Problem P2.1-nc)
\[
\max_{\tilde{V}} \quad \text{Tr}(U \tilde{V}) + c, \quad \text{s.t.} \quad \tilde{V}_{n,n} = 1, \forall n \in \{1, \ldots, N + 2\},
\]
\[
\tilde{V} \succeq 0, \quad \text{rank}(\tilde{V}) = 1.
\tag{26} \]

Dropping the rank-one constraint on \( \tilde{V} \), (26) becomes a convex semidefinite program (SDP), and can be solved straightforwardly using CVX [13]. An approximate rank-one solution \( \tilde{V}_{SDR} \) can then be obtained using Gaussian randomization. The decomposed vector \( \tilde{v}_{SDR} \) is then substituted into (24) as \( \bar{v}_0 \) to obtain a new \( U \), and the process is repeated until convergence. Note that \( F(\bar{v}) \) in (22) is bounded above, as \( R \) and \( S \) are constant matrices, and \( \| \bar{v} \|^2 = N \), which is a finite constant. Therefore, the MM algorithm is guaranteed to converge to at least a local optimum.

C. IRS Phase Shift Design: Finite Circuit Power Constraint

When \( \xi > 0 \), we note that \( \Theta \) appears in both the numerator and denominator of [13]. Therefore, FP techniques can be used to obtain a solution for \( \Theta \). The minimization of a fractional objective function \( F(x) = \frac{A(x)}{B(x)} \) over \( x \) is equivalent to
maximize its reciprocal. Hence, we can use the Dinkelbach transform [16] to rewrite the original form as (Problem P2-c)

$$\max_\theta \left| \frac{b(t)}{\| H_{Rl}^T \Theta h_{Tl} + h_{TR} \|} \left( H_{Rl}^T \Theta h_{CI} + h_{CT} \right) \right|^2,$$

s.t. $0 \leq \theta_n \leq 2\pi, \forall n \in \{1, \ldots, N\}$. (27)

Problem P2-c can then be readily solved by maximizing over $A(\Theta) - y B(\Theta)$, with the quantity $y(t + 1) = A(\Theta(t))$ updated at every iteration $t$. However, each iteration requires the MM algorithm and SDR sub-procedure outlined in the previous subsection, which incurs significant complexity. Therefore, we propose approximate solutions for the circuit constraint-limited and noise-limited regimes. In the former, where $\frac{\xi}{\eta} |b(t)|^2 | h_{Rl}^T \Theta h_{Tl} + h_{TR} |^2 \geq \gamma_{th} \sigma_R^2$, the minimized transmit power can be obtained in a similar manner as the alternating optimization algorithm in [8, Eq. (22)], where the alternating steps are performed over $w$ and $\Theta$. In the latter, where $\frac{\xi}{\eta} |b(t)|^2 | h_{Rl}^T \Theta h_{Tl} + h_{TR} |^2 \ll \gamma_{th} \sigma_R^2$, the problem simplifies to be similar to (15), and can be solved using the proposed method in Section III-B.

D. Extension to Monostatic BackCom Systems

The analysis of a monostatic IRS-aided BackCom system is a special, simpler case of the analysis for bistatic systems in the two previous subsections. Given a single antenna at the reader, all previous channel gains possessing subscript $C$ are changed to $R$. Assuming reciprocal channels, fewer unique channels are present (R-T, R-I, T-I compared to C-T, C-I, T-I, R-T, I-R), and leads to the following rearrangement of (7):

$$H_1(\Theta) = h_{Rl}^T \Theta h_{Rl} + h_{Rl}^T = H_2(\Theta)^H. \quad (28)$$

As a result, the objective function in (15) becomes

$$F(\Theta) = \| H_2(\Theta) H_2(\Theta)^H \|^2 = \| H_2(\Theta) \|^2. \quad (29)$$

Assuming semipassive tags, we can then maximize $| H_2(\Theta) |^2$, with the optimal solution for each individual phase shift being

$$\theta_n^* = \theta_{TR} - \theta_{Rl,n} - \theta_{Tl,n}, \quad (30)$$

where $\theta_n^*$ is the optimal phase of the $n$-th IRS reflector; and $\theta_{TR}, \theta_{Rl,n}$ and $\theta_{Tl,n}$ are the phases of the channels from tag to reader, the $n$-th IRS reflector to the reader, and the tag to the $n$-th IRS reflector, respectively. Due to the simpler form of the objective function in (29), the minimized transmit power expression is also simpler, with $P_{\text{opt}} = \gamma_{th} \sigma_R^2 | H_2(\Theta^*) |^{-2}$.

IV. Numerical Results

In this section, we numerically evaluate the power consumption of the IRS-aided BackCom system. We assume the CE to have $L = 4$ antennas and carrier frequency 915 MHz. The CE is located at the origin and the reader is located at $[100, 0]$, with all coordinates being in meters hereafter. We adopt the path loss model in [7, Eq. (23)], which applies to both near- and far-field IRS transmissions. The IRS is oriented towards direction $[0, 1]^T$, and is assumed to be square with square reflectors, each of width $\lambda$ without loss of generality. We consider an outdoor scenario with path loss exponent $2.2$, and for simplicity of analysis, the channel coefficients have unit magnitude with random phase drawn from a uniform distribution over $[0, 2\pi]$. Unless otherwise noted, the number of channel realizations is 100; the tag is assumed to be semi-passive with $\eta = 1$; $| b(t) | = 1$ for all tag impedances; the SNR requirement at the reader is $\gamma_{th} = 8 \text{ dB}$; and the noise power at the reader is $\sigma_R^2 = -110 \text{ dBm}$.

Fig. 2 shows the minimized transmit power of the CE as the tag is moved on a straight line between the CE and the reader. For this experiment, $N = 64$; the IRS is located at $[20, 20]$; and the tag location is between $[5, 0]$ and $[95, 0]$. In addition to solving Problem P2.1-nc using the algorithm outlined in Section III-B, we compute the suboptimal transmit power using several benchmark schemes, including: a no-IRS system; randomly generated IRS phase shifts; optimal IRS phase shifts for maximizing the received signal strength of the C-I-R link only; and optimal IRS phase shifts for maximizing the received signal strength of the T-I-R link only.

It is evident that notable power reductions are realized at all tag locations compared to the non-IRS-aided system, when the IRS phase shifts from the proposed solution are used; and the reductions are maximized when the tag is close to the IRS. For the two benchmark schemes that solely phase-align either the C-I-T link or the T-I-R link, they perform close to the MM solution when the tag is closer to the reader or the CE, respectively. This is due to the far-field nature of the respective links providing larger gains, given the tag’s location. However, when the tag is roughly halfway between the CE and reader, the proposed MM solution outperforms both benchmarks, as it selectively phase-aligns each reflector with the more favorable of the C-I-T and T-I-R links.

2The path losses for the C-I-T and T-I-R links are accounted for as constants, $\sqrt{d_{CT}}$ and $\sqrt{d_{TR}}$, and absorbed into $H_1(\Theta)$ and $H_2(\Theta)$, respectively; while the path losses for the C-T and T-R links are absorbed into $h_{CT}$ and $h_{TR}$, respectively, consistent with [4].
the non-IRS benchmark, the minimized transmit power is not symmetric with tag location, suggesting that the location of the IRS also influences the extent of power reduction.

Fig. 3 plots the CE transmit power as a function of the number of IRS reflectors. The center of the IRS is located at [20,20] and the tag is located at [20, 0]. Compared to the non-IRS-aided baseline system, it is clear that the reduction in transmit power at the source is proportional to the number of IRS reflectors. Given that passive devices such as BackCom rely completely on external powering signals for their communication, the reduction of the system power consumption is significant. For example, 5 dB power saving is achieved with a moderately small-size IRS with $N = 49$ and 10 dB power saving is achieved using an IRS with $N = 100$.

Fig. 4 reveals the transmit power reductions for a monostatic system with a single-antenna reader as a function of the reader-tag distance, with the center of the IRS located at [40,0] and oriented towards [-1 0]$. The angle of the tag relative to the reader-IRS link is varied from 0 to $. Similar levels of power saving is observed, with the largest saving occurring when the tag is collinear with, and between the reader and IRS. One may also observe that while higher transmit power is incurred as the reader-tag distance increases, the extent of power reduction also improves compared to the non-IRS benchmark. It is expected that further range increases can be realized in the case where the reader utilizes multiple antennas to perform transmit and receive beamforming.

V. CONCLUSION

In this paper, a novel IRS-aided BackCom system was introduced, where the backscatter transmission from the BackCom device is further assisted by the IRS. Accounting for multiple additional paths for reflected signals, the transmit power minimization problem at the CE was studied, through the use of the MM and SDR optimization techniques. It was shown that the addition of an IRS is able to considerably decrease the transmit power at the CE, even with a moderately small-sized IRS. In addition, the optimization framework is readily applicable to monostatic BackCom systems, resulting in similar reductions in transmit power. As an initial work into IRS-aided BackCom systems, there remains much to be studied, including the extension to multi-user scenarios.

REFERENCES

[1] J. Kimionis, A. Bletsas, and J. N. Sahalos, “Increased range bistatic scatter radio,” IEEE Trans. Commun., vol. 62, no. 3, pp. 1091–1104, Mar. 2014.
[2] V. Liu, A. Parks, V. Talla, S. Gollakota, D. Wetherall, and J. R. Smith, “Ambient backscatter: wireless communication out of thin air,” ACM SIGCOMM Comput. Commun. Rev., vol. 43, no. 4, pp. 39–50, Oct. 2013.
[3] C. Liaskos, S. Nie, A. Tsiovitiadou, A. Pitsillides, S. Ioannidis, and I. Akyildiz, “A new wireless communication paradigm through software-controlled metasurfaces,” IEEE Commun. Mag., vol. 56, no. 9, pp. 162–169, Sep. 2018.
[4] Ö. Özdogar, E. Björnson, and E. G. Larsson, “Intelligent reflecting surfaces: Physics, propagation, and pathloss modeling,” to appear in IEEE Wireless Commun. Lett., 2019.
[5] E. Basar, “Transmission through large intelligent surfaces: A new frontier in wireless communications,” in Proc. European Conf. Netw. Commun. (EuCNC), Jun. 2019, pp. 112–117.
[6] Y. Han, W. Tang, S. Jin, C. Wen, and X. Ma, “Large intelligent surface-assisted wireless communication exploiting statistical CSI,” IEEE Trans. Veh. Technol., vol. 68, no. 8, pp. 8238–8242, Aug. 2019.
[7] S. W. Ellingson, “Path loss in reconﬁgurable intelligent surface-enabled channels,” arXiv preprint arXiv:1912.06759, 2019.
[8] Q. Wu and R. Zhang, “Intelligent reﬂecting surface enhanced wireless network via joint active and passive beamforming,” IEEE Trans. Wireless Commun., vol. 18, no. 11, pp. 5394–5409, Nov. 2019.
[9] C. Pan, H. Ren, K. Wang, W. Xu, M. Elkashlan, A. Nallanathan, and L. Hanzo, “Multicell MIMO communications relying on intelligent reﬂecting surface,” arXiv preprint arXiv:1907.10864, 2019.
[10] Q. Wu and R. Zhang, “Weighted sum power maximization for intelligent reﬂecting surface aided SWIPT,” IEEE Wireless Commun. Lett., pp. 1–1, 2019.
[11] Y. Tang, G. Ma, H. Xie, J. Xu, and X. Han, “Joint transmit and reﬂective beamforming design for IRS-assisted multiuser MISO SWIPT systems,” arXiv preprint arXiv:1910.07156, 2019.
[12] W. Zhao, G. Wang, S. Atapattu, T. A. Tsiftsis, and X. Ma, “Performance analysis of large intelligent surface aided backscatter communication systems,” IEEE Wireless Commun. Lett., pp. 1–1, 2020.
[13] Z.-Q. Luo and S. Zhang, “A semidefinite relaxation scheme for multivariate quartic polynomial optimization with quadratic constraints,” SIAM J. Optim., vol. 20, no. 4, pp. 1716–1736, 2010.
[14] Z. Q. Luo and S. Zhang, “A semidefinite relaxation scheme for multivariate quartic polynomial optimization with quadratic constraints,” SIAM J. Optim., vol. 20, no. 4, pp. 1716–1736, 2010.
[15] Z. Q. Luo and S. Zhang, “A semidefinite relaxation scheme for multivariate quartic polynomial optimization with quadratic constraints,” SIAM J. Optim., vol. 20, no. 4, pp. 1716–1736, 2010.
[16] Y. Sun, P. Babu, and D. P. Palomar, “Majorization-minimization algorithms in signal processing, communications, and machine learning,” IEEE Trans. Signal Process., vol. 65, no. 3, pp. 794–816, Feb. 2017.
[17] M. Grant and S. Boyd, “CVX: Matlab software for disciplined convex programming, version 2.1,” http://cvxr.com/cvx, Mar. 2014.
[18] W. Dinkelbach, “On nonlinear fractional programming,” Manage. Sci., vol. 13, no. 7, pp. 492–498, 1967.