Intelligently Wireless Batteryless RF-Powered Reconfigurable Surface

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Abstract—This work exploits commodity, ultra-low cost, commercial radio frequency identification tags (RFID) as the elements of a reconfigurable surface. Such batteryless tags are powered and controlled by a software-defined (SDR) reader, with properly modified software, so that a source-destination link is assisted, operating at a different carrier frequency. In terms of theory, the optimal gain and corresponding best element configuration is offered, with tractable polynomial complexity (instead of exponential) in number of elements. In terms of practice, a concrete way to design and prototype a wireless, batteryless, RF-powered, reconfigurable surface is offered and a proof-of-concept is experimentally demonstrated. It is also found that even with perfect channel estimation, the weak nature of backscattered links limits the performance gains, even for large number of surface elements. Impact of channel estimation errors is also studied. Future extensions at various carrier frequencies could be directly accommodated, through simple modifications in the antenna and matching network of each RFID tag/surface element.

Index Terms—Backscatter Radio, RFID, Gen2, Reconfigurable Surface.

I. INTRODUCTION

Significant interest has been attracted recently on reconfigurable reflecting surfaces, which are viewed as a way to control the environment, with a large number of passive elements, i.e., without amplification; such surfaces are envisioned to offer a focusing effect (e.g., work in [1], [2], and references therein). A small number of experimental testbeds has recently emerged, e.g., work in [3], [4], [5]. [3] offered a 36-element array, with phase shifters helping endpoints whose line-of-sight (LoS) is blocked. [4] developed a reconfigurable intelligent surface (RIS) with 256 elements exploiting positive intrinsic-negative (PIN) diodes for 2–bit phase shifting; [5] utilized software-controlled, 2-load RF switches in groups, in order to select different surface configurations with thousands of elements, exploiting feedback from the destination. All offered testbeds so far are based on wired prototypes.

Despite the fact that RIS is a special case of bistatic backscatter radio [6], such connection is not widely known in the literature. This work exploits commodity, ultra-low cost, commercial radio frequency identification tags (RFID) as the RIS elements. Such batteryless tags are powered and controlled by a software-defined (SDR) reader, operating at carrier frequency $f_2$, with properly modified software, so that a source-destination link, operating at a different carrier frequency $f_1$, is assisted (Fig. 1).

It is found for the first time in the literature (to the best of our knowledge): a) the optimal gain and corresponding best RIS element configuration, with tractable polynomial complexity (instead of exponential) in number of elements and b) the way to design and prototype a wireless, batteryless, RF-powered RIS, with commodity RFID tags. Future extensions at various carrier frequencies could be accommodated, through simple modifications in the antenna and matching network of each RFID tag.

Notation: $0_N$ denotes the all-zeros vector. The phase of complex number $z$ is denoted as $\angle z$, while $\Re\{z\}$ denotes the real part of $z$. The distribution of a proper complex Gaussian $N \times 1$ vector $x$ with mean $\mu$ and covariance matrix $\Sigma$ is denoted by $\mathcal{CN}(\mu, \Sigma) \triangleq \frac{1}{\det(\Sigma)^{1/2}} e^{-\frac{1}{2}(x-\mu)^{\mathsf{T}} \Sigma^{-1} (x-\mu)}$; the special case of a circularly symmetric complex Gaussian $N \times 1$ vector corresponds by definition to $\mathcal{CN}(0_N, \Sigma)$; expectation of function $g(\cdot)$ of random variable $x$ is denoted by $\mathbb{E}[g(x)]$.

II. SYSTEM MODEL

A. Channel Model

A source-destination link is assisted by an array of $M$ tags/RIS elements. The following large-scale channel pathloss model is adopted [7]:

$$L_X \propto \left( \frac{\lambda}{4\pi d_0} \right)^2 \left( \frac{d_0^4}{d_X} \right)^{\nu_X}, \quad (1)$$

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where \( X \in \{ \text{SD}, \text{ST} m, T_mD \} \) denotes the source-to-destination, source-to-tag \( m \) and tag \( m \)-to-destination, respectively; \( \lambda \) is the carrier wavelength, \( d_0^m \) is a reference distance, \( v_X \) is the path-loss exponent and \( d_X \) is the distance for link \( X \).

Flat fading is assumed; complex channel coefficient \( h_{\text{SD}} \), \( h_{\text{ST} m} \) and \( h_{T_mD} \) denotes the baseband channel coefficients for the source-destination, source-tag and tag-reader link, respectively. Due to strong line-of-sight (LoS) signals present in the problem, small-scale Rice flat fading channel model \( 7 \) is mainly adopted \[ h_{T_mD} \sim \mathcal{CN} \left( \sqrt{\frac{\nu_{T_mD}}{\kappa_{T_mD} + 1}} \sigma_{T_mD}, \frac{\sigma_{T_mD}^2}{\kappa_{T_mD} + 1} \right), \] (2)

where \( h_{T_mD} \triangleq |h_{T_mD}| e^{-j\phi_{T_mD}} \), \( k_{T_mD} \) is the power ratio between the deterministic LoS component and the scattering components and \( \mathbb{E}[|h_{T_mD}|^2] = \sigma_{T_mD}^2 \) is the average power of the scattering components. For link budget normalization purposes, \( \sigma_{T_mD} = 1 \) will be also assumed (other values could be easily accommodated into the large-scale, average coefficients). Similar notation and assumptions hold for \( h_{\text{ST} m}, m \in \{1, 2, \ldots, M\} \) and \( h_{\text{SD}} \). It is noted that for \( \kappa = 0 \), Rice is simplified to Rayleigh fading.

Quasi-static block fading is assumed, i.e., the channel remains constant for \( L_c \) (source-destination link) symbols and changes independently between channel coherence time periods. Channel coefficients \( h_{\text{SD}}, \{h_{\text{ST} m}\}, \{h_{T_mD}\}, m \in \{1, 2, \ldots, M\} \) are assumed independent in the numerical results. Furthermore, the following notation is also adopted:

\[ h_m = h_{\text{ST} m} h_{T_mD} = |h_{\text{ST} m} h_{T_mD}| e^{-j\phi_m}, m \in \{1, 2, \ldots, M\}, \]
\[ h_0 = h_{\text{SD}}, m = 0, \] (3)

### B. Signal Model

The baseband source message \( m(t) \) is given by:

\[ c(t) = \sqrt{2P} m(t) \] (4)

where \( \mathbb{E}[|m(t)|^2] = 1 \). Different normalization could be incorporated into the large-scale coefficients. The baseband complex equivalent of the scattered waveform from tag \( m \) is given by \[ 6 \] :

\[ u_m(t) = \sqrt{\eta} h_{\text{ST} m} [A_s - \Gamma_m(t)] h_{T_mD} c(t), \]
\[ \Gamma_m(t) \in \{ \Gamma_1, \Gamma_2, \ldots, \Gamma_K \}, \] (5)

where \( \Gamma_m(t) \) stands for the modified (complex) reflection coefficient for tag \( m \), assuming that the tag can terminate its antenna between \( K \) loads and \( \eta \) models the (limited) tag power scattering efficiency. It is noted that for passive (amplification-free) tags, \( \Gamma_k \leq 1 \), while for commercial RFID tags, \( K = 2 \). Parameter \( A_s \) stands for the load-independent structural mode that solely depends on tag’s antenna \[ 8 \], commonly overlooked in the literature; \( A_s = 0 \) only for minimum scattering antennas, i.e., antennas that do not reflect anything when terminated at open (i.e., infinite) load.

The received demodulated complex baseband signal at the destination is given by the superposition of the source and all tags’ backscattered signals propagated through wireless channels \( h_{\text{SD}} \) and \( \{h_{T_mD}\} \), respectively:

\[ y(t) \triangleq \sqrt{L_{\text{SD}}} h_{\text{SD}} c(t) + \sum_{m=1}^{M} \sqrt{L_{T_mD}} h_{T_mD} u_m(t) + n(t) \]
\[ = \sqrt{L_{\text{SD}}} h_{\text{SD}} c(t) + n(t) \]
\[ + \sum_{m=1}^{M} \sqrt{\eta} h_{\text{ST} m} h_{T_mD} [A_s - \Gamma_m(t)] c(t) \]
\[ = \sqrt{2P} \left[ \sqrt{g_0} h_0 + \sum_{m=1}^{M} \sqrt{g_m} h_m \mathcal{Y}_m(t) \right] m(t) + n(t), \] (7)

where \( n(t) \) is the thermal noise at the receiver and

\[ g_0 = L_{\text{SD}}, \]
\[ g_m = \eta L_{\text{ST} m} L_{T_mD} \mathbb{E} \left[ |A_s - \Gamma_m(t)|^2 \right], \]
\[ \mathcal{Y}_m(t) = \frac{A_s - \Gamma_m(t)}{\sqrt{\mathbb{E} \left[ |A_s - \Gamma_m(t)|^2 \right]}} \] (8)

\[ y_m [\Gamma_m(t)] \triangleq \sqrt{g_m} h_m \mathcal{Y}_m(t). \] (9)

Notice that \( \mathbb{E}[|h_0|^2] = \mathbb{E}[|h_m|^2] = \mathbb{E}[|\mathcal{Y}_m|^2] = 1 \) since \( \mathbb{E}[|h_m|^2] = \mathbb{E}[|h_{\text{ST} m}|^2] \mathbb{E}[|h_{T_mD}|^2] \), due to the followed assumptions. It is also noted that \( \mathbb{E} \left[ |A_s - \Gamma_m(t)|^2 \right] = \left( 1/K \right) \sum_{k=1}^{K} |A_s - \Gamma_k|^2. \)

The above model is valid when coupling among the tags is negligible, i.e., the tags are separated by distance at least equal to \( \lambda/2 \). Additive thermal noise \( n(t) \) is modelled by a complex, circularly symmetric, additive Gaussian noise process with \( \mathbb{E}[|n(t)|^2] = N_0 B \), where \( B \) stands for receiver’s bandwidth \[ 4 \].

### III. Optimal Gain

RIS targets at SNR improvement, by controlling the environment, through proper selection of the reflection coefficient at each element. According to Eq. (7), the following instantaneous power maximization problem is formulated:

\[ \max_{\{\mathcal{Y}_m(t)\}} \left| \sqrt{g_0} h_0 + \sum_{m=1}^{M} \sqrt{g_m} h_m \mathcal{Y}_m(t) \right|^2 \]
\[ = \max_{\{\Gamma_m(t)\}} \left| y_0 + \sum_{m=1}^{M} y_m [\Gamma_m(t)] \right|^2 \] (10)

which cannot be solved with exhaustive search, since each RIS element (among \( M \) elements) can select among \( K \) loads, i.e., \( \Gamma_m(t) \in \{ \Gamma_1, \Gamma_2, \ldots, \Gamma_K \} \), and thus, there are \( K^M \) possible load configurations. Even for \( K = 2 \) and \( M = 100 \), exhaustive search among \( 2^{100} \) load configurations is not an option.

\[ 2 \] The complex channel is the superposition of \( \sqrt{\frac{\nu_{T_mD}}{\kappa_{T_mD} + 1}} \sigma_{T_mD} e^{j\theta} + \mathcal{CN} \left( 0, \frac{\sigma_{T_mD}^2}{\kappa_{T_mD} + 1} \right) \) with random \( \theta \).
The problem above is similar to noncoherent sequence detection of orthogonally-modulated sequences, solved with polynomial complexity in [9]. The trick is to introduce an auxiliary scalar variable $\phi$ into the problem of Eq. (14):

$$\max_{\{\Gamma_m(t)\}} \max_{\phi \in [0, 2\pi]} \Re \left\{ e^{-j\phi} \left( y_0 + \sum_{m=1}^{M} y_m \left[ \Gamma_m(t) \right] \right) \right\} =$$

$$\max_{\phi \in [0, 2\pi]} \max_{\{\Gamma_m(t)\}} \left( \Re \left\{ e^{-j\phi} y_0 \right\} + \sum_{m=1}^{M} \Re \left\{ e^{-j\phi} y_m \left[ \Gamma_m(t) \right] \right\} \right)$$

(15)

A. $K = 2$ Loads

For a given point $\phi \in [0, 2\pi)$, the innermost maximization in Eq. (15) is separable for each $\Gamma_m(t)$ and hence, splits into independent maximizations for any $m = 1, 2, \ldots, M$:

$$\hat{\Gamma}_m(t) = \arg \max_{\Gamma_m(t) \in \{\Gamma_1, \Gamma_2\}} \Re \left\{ e^{-j\phi} y_m \left[ \Gamma_m(t) \right] \right\}$$

$$\Leftrightarrow \Re \left\{ e^{-j\phi} y_m \left[ \Gamma_1 \right] \right\} \overset{\hat{\Gamma}_m(t) = \Gamma_1}{\gtrless} \Re \left\{ e^{-j\phi} y_m \left[ \Gamma_2 \right] \right\}$$

$$\Leftrightarrow \Re \left\{ e^{-j\phi} (y_m \left[ \Gamma_1 \right] - y_m \left[ \Gamma_2 \right]) \right\} \overset{\hat{\Gamma}_m(t) = \Gamma_1}{\gtrless} 0$$

$$\Leftrightarrow \cos \left( \phi - \frac{y_m \left[ \Gamma_1 \right] - y_m \left[ \Gamma_2 \right]}{2} \right) \overset{\hat{\Gamma}_m(t) = \Gamma_1}{\gtrless} 0$$

(16)

Given the relation in Eq. (15), the optimal load sequence $\hat{\Gamma}_{o p t}$ can be found by varying $\phi$ from 0 to $2\pi$. It is further noticed that, as $\phi$ scans $[0, 2\pi)$, the decision $\hat{\Gamma}_m(t)$ changes, according to Eq. (16), only when:

$$\cos \left( \phi - \frac{y_m \left[ \Gamma_1 \right] - y_m \left[ \Gamma_2 \right]}{2} \right) = 0$$

$$\Leftrightarrow \phi = \pm \frac{\pi}{2} + \frac{y_m \left[ \Gamma_1 \right] - y_m \left[ \Gamma_2 \right]}{2} \left( \text{mod} \ 2\pi \right).$$

(17)

Therefore, the sequence $\hat{\Gamma} = \left[ \hat{\Gamma}_1(t), \hat{\Gamma}_2(t), \ldots, \hat{\Gamma}_M(t) \right]^T$ changes only at $\left( \phi_{m}^{(1)}, \phi_{1}^{(2)}, \phi_{2}^{(2)}, \ldots, \phi_{M}^{(1)}, \phi_{M}^{(2)} \right)$. For the remaining part of this section, we assume that the above $2M$ points are distinct and nonzero, i.e., $\phi_m^{(j)} \neq \phi_i^{(k)}$ and $\phi_m^{(j)} \neq 0$, for any $j, k, \in \{1, 2\}$ and $m, l, \in \{1, 2, \ldots, M\}$ with $m \neq l$. There is a case where the above assumption does not hold, examined in [9]. If the above points are sorted in ascending order, i.e.,

$$(\theta_1, \theta_2, \cdots, \theta_{2M}) =$$

$$= \text{sort} \left( \phi_{1}^{(1)}, \phi_{1}^{(2)}, \phi_{2}^{(2)}, \ldots, \phi_{M}^{(1)}, \phi_{M}^{(2)} \right),$$

(18)

then the decision $\hat{\Gamma}$ will remain constant in each one of the 2$M$ + 1 intervals $[\theta_1, \theta_{i+1})$, $i \in \{0, 1, \ldots, 2M\}$, with $\theta_0 = 0$ and $\theta_{2M+1} = 2\pi$. The goal is the identification of the 2$M$ + 1 sequences that correspond to these intervals, one of which gives the optimal $\hat{\Gamma}_{o p t}$, i.e., the one that offers the maximum power; thus, the quality of each sequence is calculated with the norm metric of Eq. (14), which explicitly includes the direct channel $h_0$. Based on the above, the sorting operation in Eq. (15) is dominant in terms of computational cost, which is $O(M \log M)$ and not $2^M$.

B. $K > 2$ Loads

The method described above can be generalized to $K > 2$ loads, i.e., $\Gamma_m(t)$ belongs in $\{\Gamma_1, \Gamma_2, \ldots, \Gamma_K\}$. The solution is given by selecting the largest value of $\Re \left\{ e^{-j\phi} y_m \left[ \Gamma_k \right] \right\}$ among all $k \in \{1, 2, \ldots, K\}$, which results in testing $2M \times (K - 1)$ changes of $\phi$ and as a result, same number of sequence changes and not $2M \times (\frac{K}{2})$, as one would expect; the rest of the steps are exactly the same as in $K = 2$. Formal proof and details can be found in [9], omitted due to space constraints. Notice that the norm metric for the quality of each sequence must include $h_0$. The complexity of the algorithm is again $O(M \log M)$ for $M > K$ and not $K^M$.

Fig. 2 left depicts $K = 21$ (complex) reflection coefficients, corresponding to the 21 loads, offered through a varactor at each tag, as experimentally tested in [10]. It can be shown that these reflection coefficients span more than $120^\circ$. Fig. 2 right shows $A_s - \Gamma_k, k \in \{1, 2, \ldots, K\}$, which incorporates the contribution of the structural mode of each RIS element antenna, typically overlooked in the literature; it can be seen that the span of $A_s - \Gamma_k$ in degrees is much smaller, in the order of $60^\circ$.

IV. WIRELESS BATTERYLESS IMPLEMENTATION

The RFID-based RIS is controlled by the RFID reader at carrier frequency $f_2$, through commands such as Select, Query and ACK, explained below. In the commercial RFID standard (EPC Gen2), a framed slotted Aloha (FSA) protocol is used, so that tags backscatter one at a time, without collision. For RIS purposes, the opposite is needed, i.e., tags must be forced to backscatter in carefully selected groups.

In Gen2, the reader initiates the start of a frame with a Query command, which contains the number of slots. Ideally, the number of slots should be equal to the number of tags to be inventoried. If the reader advertises number of slots equal to 1, then all tags in the vicinity of the reader will respond. That was the approach followed in this work. Then, each tag responds with a random 16-bit number, namely the RN16 message, which is in principle different among the competing tags. If the reader correctly decodes that message then it will send the ACK/command, containing the RN16 sent by the
The RN16 is preceded by a 6-bit Preamble sequence, which does not follow the line coding rules and is the same for all tags. As a result, using the Preamble one could measure the effect of a specific RIS configuration (where configuration means the set of tags that change their reflection coefficients, among the total number of RIS tags). Since the random 16-bit sequence of the RN16 is different for each tag, the signal (or the effective channel) during RN16 could be higher or lower depending on which tags are selected. T

![Inventory round without RN16 collision](inventory-round.png)  ![Inventory round with RN16 collision](inventory-round.png)

(a) No RN16 collision. (b) RN16 collision.

Fig. 3: Conventional Gen2 RFID operation: each tag responds to a Query command with a random RN16 message; if it is correctly acknowledged it will respond with its EPC.

The Select command is asserting or deasserting the SL flag of the specified tags, sequentially. This is shown in Fig. 4.

![RN16 at source-destination frequency (870 MHz)](rn16-source-destination.png)

Fig. 4: Superposition of Preamble+RN16 from multiple tags.

The RN16 is set to \( t_c \), where \( t_c \) is the channel coherence time. Different configurations offer different effective channel and hence, different maximum signal power. Those peaks occur at the Preamble+RN16 tag response, as explained in Fig. 5. It is noted that the measurements were conducted in a static indoor environment, with channel coherence time spanning several hundreds of msec (and thus, different peaks are due to different tag configurations and not changes in the channel). The time between the last RN16 from one configuration and the first RN16 from the next is specified as \( t_c \). More specifically, by varying the backscatter link frequency (BLF) parameter, \( t_c \) changes accordingly; for example, if 50 active tags are selected (among a population of \( M \)), then \( t_c \in [43 \text{ ms}, 116 \text{ ms}] \). In Fig. 6, \( t_c = 116 \text{ ms} \). Future RIS-friendly modifications of Gen2 could further reduce this parameter.

For RIS purposes, a specific configuration is set by having the reader issuing successive Select commands and asserting the SL flag of the specified tags, sequentially. This is shown in Fig. 5. First, a continuous wave (CW) energizes the tags and then the Select commands are issued. Afterwards, we can observe 10 inventory rounds (Query-RN16-ACK), where the peaks are the superposition of the RN16 from the selected tags. Finally, to advance to a new configuration, another Select is sent but with Action parameter equal to \( \{0, 0, 0\} \) and a non-matching mask to any tag.

![Reader-Tag communication during a configuration](reader-tag-communication.png)

Fig. 5: Reader-Tag communication during a configuration.

![Power at Destination during different configurations](power-destination.png)

Fig. 6: Power at Destination during different configurations.

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4The reader directs the tags about the line code they are going to use.
The experimental setup is depicted in Fig. 7 with $M = 100$ commodity Zebra Z-Perform 1500T Gen2 RFID tags, separated by $d_x = 10$ cm and $d_y = 5$ cm. The source-destination link operates at $f_1 = 870$ MHz, while the software-defined radio (SDR) reader operates at $f_2 = 868$ MHz, with zero interference among the two channels. Modifications in the Gen2 reader software stack from [11] were conducted to enable RIS functionality. Two commodity USRP N200 software-defined radios (SDR), each equipped with a RFX900 daughterboard, were utilized, both connected to a commodity laptop through a switch. The antennas used at the reader and source were circularly-polarized. It is emphasized that Figs. 3, 4, 5, 6 are offered from the SDR receiver at the destination.

V. NUMERICAL RESULTS

A. Simulations

Figs. 8 and 9 are generated with typical indoor static conditions in mind: $\kappa_{ST_m} = \kappa_{TD_m} = \kappa_{SD} = 8$, $\eta = 10\%$, $v_X = 3$, $d_y^B = d_{SD} = 3$ m, $f_2 = 870$ MHz, $P = 5$ dBm (3.16 mW), and $10$ dB relative end-2-end antenna gain for the backscattered links compared to direct link, assuming that the source and destination antennas point towards the RIS (in order to assist its operation). The same parameters hold for Fig. 10 unless otherwise noted.

Fig. 8 offers the average power improvement due to RIS and direct link operation, compared to power of the direct link communication. Several instantiation of the channels are generated, optimal configuration for $K = 2$ or $K = 21$ is found, based on the analysis of Section III-B and average values are reported. The SD link is parallel to RIS and distance from surface $(d_{RIS-SD})$ is varied. It is observed that as the distance of the SD link from the RIS increases, performance gains to optimal operation of RIS decrease. That is due to the fact that backscattered links (through RIS) become weaker, as the SD link is farther away from the surface. This finding suggests that amplification at each RIS element (i.e., $|V_k| > 1$) is needed. Finally, it is observed that using $K = 21$ loads for each element, despite the small angle span of the induced backscattered signals (explained in Sec. III-B), offers power gain in the order of $1.3$ additional dB, compared to $K = 2$.

Fig. 9 offers average power improvement as a function of $M$, for two setups: Fig. 9 left sets $d_x = 0.1$ m and $d_y = 0.05$ m (setup (a)), while Fig. 9 right sets spacing between adjacent tags equal to $\lambda/2$ (setup (b)). For both setups, $d_{SD} = 3$ m, $d_{RIS-SD} = 1$ m, while setup (a) ignores possible coupling between adjacent tags. It can be noticed that performance in (b) reaches a plateau with a faster rate than in (a), as a consequence of the larger element spacing. This finding suggests that even with perfect channel estimation, the weak nature of backscattered links limits the performance gains, even for large number of tags/RIS elements.

Finally, Fig. 10 quantifies the impact of channel estimation error $\hat{h}_m$ on the estimation of $h_m$, with $\hat{h}_m$ being the channel estimate, i.e., $h_m = \hat{h}_m + \hat{h}_m$, assuming minimum channel estimation error (MMSE) estimators, with $\mathbb{E}[|\hat{h}_m|^2] = \text{MMSE}_m$ and expressions for MMSE, $m \in \{0, 1, 2, \ldots, M\}$ from the multi-antenna literature [12, 13]. Rayleigh fading is assumed ($\gamma_{ST_m} = \gamma_{TD_m} = \gamma_{SD} = 0$), with $v_X = 3$, $d_{RIS-SD} = 8$ m, $d_y^B = 3$ m, $d_{SD} = 15$ m and $B = 48$ MHz. The optimal configuration is found with estimated channels $\{\hat{h}_m\}$,
while the effective channel and gains are found based on the true channels and the offered configurations. Channel coherence time in number of symbols is set to $L_c = 24 \times 10^5$, corresponding to channel coherence time of 100 ms and SD link at 48 Mbps with QPSK modulation. If $\alpha$ is the percentage of $L_c$ devoted to pilot transmission for channel estimation then $\alpha L_c > M$, since the number of pilot symbols cannot be less than the number of unknown channels; in this plot $\alpha$ varies from 1% to 11% as $M$ increases. It is found that channel estimation error for large number of elements $M$ has important, non-negligible impact.

B. Experiments

Fig. [11] offers the received power at the destination, operating @ 870 MHz, for 2 different setups of Fig. [7]. With the reader on (@ 868 MHz), both the RIS and the SD link contribute to the received power (@ 870 MHz). With the reader switched off, the received power at the destination is measured and depicted. Clearly, one setup offers constructive and another destructive RIS operation. The source-RIS distance was denoted as $d_{\text{RIS-S}}$ and the destination-RIS as $d_{\text{RIS-D}}$, measured from the RIS center; $\phi_{\text{SD}}$ is the angle between the direction of the destination antenna and the SD link. For the constructive setup, $d_{\text{RIS-S}} = 2.5$ m, $d_{\text{RIS-D}} = 1.1$ m, $\phi_{\text{SD}} = 100^\circ$ and $d_{\text{SD}} = 2.1$ m. For the destructive setup, $d_{\text{RIS-S}} = 2.1$ m, $d_{\text{RIS-D}} = 1.1$ m, $\phi_{\text{SD}} = 80^\circ$, and $d_{\text{SD}} = 2$ m.

Fig. [12] shows the maximum power improvement achieved as a function of number of configurations tested, for variable number of activated tags (denoted by $\mu$). The setup corresponds to $d_{\text{RIS-S}} = 2.2$ m, $d_{\text{RIS-D}} = 0.9$ m, $\phi_{\text{SD}} = 145^\circ$ and $d_{\text{SD}} = 1.4$ m. For a given configuration and $\mu$, the value reported is over three repetitions. Increasing the number of configurations tested results to an increased maximum power value. Also, when $\mu$ is increased, the maximum gain is increased for a given number of configurations. The increase of the power improvement in Fig. [12] at 6.5 dB versus the 2.5 dB of Fig. [11] left is due to the change in the antenna gain for different $\phi_{\text{SD}}$, since the destination antenna is directive, affecting the SD link’s contribution.

REFERENCES

[1] C. Huang, S. Hu, G. C. Alexandropoulos, A. Zappone, C. Yuen, R. Zhang, M. D. Renzo, and M. Debbah, “Holographic MIMO Surfaces for 6G Wireless Networks: Opportunities, Challenges, and Trends,” IEEE Trans. Wireless Commun., vol. 27, no. 5, pp. 118–125, Oct. 2020.

[2] E. Björnson, Ozgecan Özdogan, and E. G. Larsson, “Reconfigurable Intelligent Surfaces: Three Myths and Two Critical Questions,” IEEE Commun. Mag., vol. 58, no. 12, pp. 90–96, Dec. 2020.

[3] Z. Li, Y. Xie, L. Shangguan, R. I. Zelaya, J. Gummesson, W. Hu, and K. Jamieson, “Towards Programming the Radio Environment with Large Arrays of Inexpensive Antennas,” in 16th USENIX Symposium on Networked Systems Design and Implementation (NSDI 19), Boston, MA, Feb. 2019, pp. 285–300.

[4] L. Dai, B. Wang, M. Wang, X. Yang, J. Tan, S. Bi, S. Xu, F. Yang, Z. Chen, M. D. Renzo, and L. Hanzo, “Reconfigurable Intelligent Surface-Based Wireless Communications: Antenna Design, Prototyping, and Experimental Results,” IEEE Access, vol. 8, pp. 45 913–45 923, Mar. 2020.

[5] V. Arun and H. Balakrishnan, “RiFocus: Beamforming Using Thousands of Passive Antennas,” in 17th USENIX Symposium on Networked Systems Design and Implementation (NSDI 20), Santa Clara, CA, Feb. 2020, pp. 1047–1061.

[6] 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.

[7] A. Goldsmith, Wireless Communications. New York, NY, USA: Cambridge University Press, 2005.

[8] A. Bletsas, A. G. Dimitriou, and J. N. Sahalos, “Improving Backscatter Radio Tag Efficiency,” IEEE Trans. Microwave Theory Tech., vol. 58, no. 6, pp. 1502–1509, Jun. 2010.

[9] P. N. Alevizos, Y. Fountzoulas, G. N. Karystinos, and A. Bletsas, “Log-Linear-Complexity GLRT-Optimal Noncoherent Sequence Detection for Orthogonal and RFID-Oriented Modulations,” IEEE Trans. Commun., vol. 64, no. 4, pp. 1600–1612, Apr. 2016.

[10] G. Vougioukas, A. Bletsas, and J. N. Sahalos, “Instantaneous, Zero-Feedback Fading Mitigation with Simple Backscatter Radio Tags,” IEEE Journal of Radio Frequency Identification, Oct. 2020, early access.

[11] N. Kargas, F. Mavromatis, and A. Bletsas, “Fully-Coherent Reader With Commodity SDR for Gen2 FM0 and Computational RFID,” IEEE Wireless Commun. Lett., vol. 4, no. 6, pp. 617–620, Dec. 2015.

[12] B. Hassibi and B. Hochwald, “How much training is needed in multiple-antenna wireless links?” IEEE Trans. Inform. Theory, vol. 49, no. 4, pp. 951–963, Apr. 2003.

[13] A. Lozano, R. W. Heath, and J. G. Andrews, “Fundamental Limits of Cooperation,” IEEE Trans. Inform. Theory, vol. 59, no. 9, pp. 5213–5226, Sep. 2013.