

Differential Reflecting Modulation for Reconfigurable Intelligent Surface Based Communications

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Abstract—Reconfigurable intelligent surface (RIS) based communications have emerged as a new paradigm. This letter proposes a differential reflecting modulation (DRM) scheme for RIS based communication systems. In DRM, information bits are jointly carried by the activation permutations of the reflecting patterns and the phases of the transmitted signals, leading to that DRM can work without any channel state information (CSI) at the transmitter, RIS or receiver. In other words, DRM can release the intricate and resource-consuming channel estimation in the transmission process. Simulation results show that the proposed DRM pays acceptable SNR penalty compared to non-differential modulation scheme with coherent detection.

Index Terms—Reconfigurable intelligent surface, differential reflecting modulation (DRM), differential detection

I. INTRODUCTION

RECONFIGURABLE intelligent surface (RIS) is a key enabler for configuring the favorable wireless communication environment [1], [2]. The research on RIS-based information transfer has attracted a lot of interest. As stated in [3], these studies could be classified into four categories: RIS-aided communications (RIS-C), RIS-based backscatter communications (RIS-BC), RIS-based spatial modulation (RIS-SM), passive beamforming and information transfer (PBIT).

The works belong to RIS-C only activate the reflection function aiming to maximize system performance regarding to various criteria. For example, Huang *et al* in [4] jointly designed the phase shifts at RIS and the power allocation at the base station (BS) to maximize the energy efficiency; Ye *et al* in [5] improved the symbol error rate performance of RIS-C systems by adjusting the precoding matrix at BS and reflection coefficients at RIS; Guan *et al* showed that the physical layer security in the RIS-assisted system can be enhanced by incorporating artificial noise or jamming in [6].

The works researched the information modulation role of RIS is classified in to RIS-BC, which is investigated in works [7]–[9] by Tang *et al*. They take advantage of the concept of the programmable metasurface and designed reflection

coefficient controllable metasurface based transmitter enabling phase modulation in [7], RF chain-free transmitter and space-down-conversion receiver in [8], and a transmitter without filter wideband mixer and power amplifier in [10]. The RIS designed in works belonging to RIS-SM combined with the spatial modulation technology, which can both reflects and carries information. Preliminary work on RIS-BC was undertaken by Basar *et al* in [11], who proposed RIS-based access point to assist the transmission of unmodulated carrier.

Combing index modulation techniques with RIS, an RIS-based space shift keying scheme was shown to enables highly reliable transmission with unconventionally high energy efficiency in [12]. More recently, Gopi *et al* in [13] designed three RIS-based architectures for beam-index modulation in millimeter wave communication to circumvent the line-of-sight blockage of millimeter wave frequencies.

In PBIT, RIS not only helps the information delivery from the transmitter to the receiver, but also delivers its own information in [14]. In this work, a passive beamforming method was developed to improve the average receive signal-to-noise ratio and a two-step approach was established to retrieve the information from both the transmitter and the RIS. [3] has made a comprehensive comparison among these four categories and proposed an optimized reflecting modulation (RM) schemes that outperforms all existing RIS-based information transfers.

So far, however, almost all literature needs to access the instantaneous/statistic channel state information (CSI). In RIS-based communication systems, the channel estimations involving the direct link, transmitter-RIS, and RIS-receiver links are intricate tasks. This is because RIS units has no baseband signal precessing capabilities. Although some literature solves this critical channel estimation problem through different methods [10], it still consumes a lot of time and resources [15]–[17]. In this letter, we propose a novel differential reflecting modulation (RIS) scheme, which completely bypasses any CSI at the transmitter, RIS or receiver.

II. SYSTEM MODEL

In this letter, a RIS-assisted (N_r, N, r) single-input multiple-output (SIMO) communication system model as illustrated in Fig. 1, where the transmitter is equipped with one antenna; N_r denotes the numbers of receiver antennas; N stands for the number of reflecting units on the surface; and r stands for the

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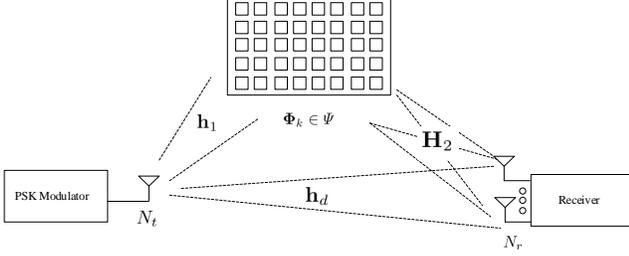


Fig. 1. A RIS-assisted (N_r, N, r) SIMO communication system.

total transmission rate. In the system, the transmitter sends M -ary phase shift-keying (PSK) modulated signals. Such system setups can model the communication between a power- and size- constrained transmitter and a powerful receiver, such as the uplink communication in cellular systems.

Let $\mathbf{h}_1 \in \mathbb{C}^{N \times 1}$, $\mathbf{H}_2 \in \mathbb{C}^{N_r \times N}$, $\mathbf{h}_d \in \mathbb{C}^{N_r \times 1}$ denote the transmitter-RIS channel vector, the RIS-receiver channel matrix, and the direct link channel vector. We assume that the channels follows Rayleigh fading and stay constant during the transmission. It is worth noting that Rayleigh fading channels are chosen as an example to demonstrate the performance and the proposed DRM scheme are also applicable to other channel models. Moreover, we assume that a total of K reflecting pattern candidates are employed for transmission, which are included in a set $\Psi = \{\Phi_1, \Phi_2, \dots, \Phi_K\}$. Each reflecting pattern can be mathematically expressed as a diagonal matrix as $\Phi_k \in \mathbb{C}^{N \times N}$. The diagonal elements of Φ_k can be expressed as $(\Phi_k)_{nn} = \beta_n^{(k)} \exp(j\theta_n^{(k)})$, $1 \leq n \leq N$, where $\beta_n^{(k)} = \{0, 1\}$ indicates the ON/OFF states of the n th reflecting unit and $\theta_n^{(k)}$ represents the phase shift angle at the n -th reflecting unit when the k th reflecting pattern Φ_k is activated.

III. DIFFERENTIAL REFLECTING MODULATION

A. Differential Encoding Scheme

During the transmission, a frame is divided into blocks with each consists of K symbol time slots. At the t th block, a total of $r = \log_2 \lfloor K! \rfloor + K \log_2 M$ bits are delivered. As shown in Fig. 2, the first $r_1 = \log_2 \lfloor K! \rfloor$ bits are mapped to a $K \times K$ permutation matrix \mathbf{Z}_t . To demonstrate the mapping process, an example with 3 available reflecting patterns is listed in Table I. In the demonstrated example, 4 permutation matrices are chosen from all $3! = 6$ feasible candidates for carrying 2 bits. It should be mentioned that different choices will result different performance. The rest $r_2 = K \log_2 M$ bits are mapped to K M -PSK symbols s_k ($1 \leq k \leq K$), which are transmitted during the K slots. By stacking all K symbols in a vector $\mathbf{s}_t = [s_1, s_2, \dots, s_K]$ and defining $\mathbf{S}_t = \text{diag}(\mathbf{s}_t) \in \mathbb{C}^{K \times K}$, we introduce an information-carrying matrix $\mathbf{X}_t \in \mathbb{C}^{K \times K}$ given by

$$\mathbf{X}_t = \mathbf{Z}_t \mathbf{S}_t. \quad (1)$$

Then, a new matrix $\mathbf{V}_t \in \mathbb{C}^{K \times K}$ can be generated after the differential encoding as

$$\mathbf{V}_t = \mathbf{V}_{t-1} \mathbf{X}_t, \quad (2)$$

TABLE I
A BIT ENCODING EXAMPLE

Bits	\mathbf{Z}_t
00	$\begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix}$
01	$\begin{bmatrix} 1 & 0 & 0 \\ 0 & 0 & 1 \\ 0 & 1 & 0 \end{bmatrix}$
10	$\begin{bmatrix} 0 & 1 & 0 \\ 1 & 0 & 0 \\ 0 & 0 & 1 \end{bmatrix}$
11	$\begin{bmatrix} 0 & 0 & 1 \\ 1 & 0 & 0 \\ 0 & 1 & 0 \end{bmatrix}$

where \mathbf{V}_{t-1} is the matrix generated in the former block and $\mathbf{V}_0 \in \mathbb{C}^{K \times K}$ is an identity matrix for initialization. Based on the definition of \mathbf{V}_t , it is easily verified that \mathbf{V}_t is a multiplication of a permutation matrix and a diagonal matrix as

$$\mathbf{V}_t = \tilde{\mathbf{Z}}_t \tilde{\mathbf{S}}_t, \quad (3)$$

where $\tilde{\mathbf{Z}}_t \in \mathbb{C}^{K \times K}$ denotes a permutation matrix and $\tilde{\mathbf{S}}_t \in \mathbb{C}^{K \times K}$ repents a diagonal matrix whose symbols are chosen from M -PSK symbol set, which is denoted by \mathcal{S}_M . This is because

$$\mathbf{V}_t = \mathbf{V}_0 \mathbf{Z}_1 \mathbf{S}_1 \mathbf{Z}_2 \mathbf{S}_2 \dots \mathbf{Z}_t \mathbf{S}_t. \quad (4)$$

As shown in Fig. 2, next step is to activate the reflecting patterns and modulate the phases of transmit signals according to \mathbf{V}_t during the t th block, which will be given in the following subsection.

B. Signal Transmission

Let $\mathbf{v}_k^{(t)} \in \mathbb{C}^{K \times 1}$ being k -th column vector of \mathbf{V}_t and according to the form of \mathbf{V}_t in (3), \mathbf{v}_k can be expressed as

$$\mathbf{v}_k^{(t)} = \mathbf{e}_i s_l, \quad (5)$$

where $\mathbf{e}_i \in \mathbb{C}^{K \times 1}$ represents the i th basis vector with i th elements being nonzero; and s_l is the nonzero element of $\mathbf{v}_k^{(t)}$ at the i th position, which is an M -PSK symbol. Then, according to $\mathbf{v}_k^{(t)}$, RIS activates the i th reflecting pattern Φ_i and the transmitter sends the M -PSK symbol s_l at the k -th slot of the t -th block. Thus, the received signal $\mathbf{y}_k^{(t)} \in \mathbb{C}^{N_r \times 1}$ in the k -th slot of the t -th block can be written as

$$\mathbf{y}_k^{(t)} = (\mathbf{h}_d + \mathbf{H}_2 \Phi_i \mathbf{h}_1) s_l + \mathbf{n}_k^{(t)}, \quad (6)$$

where \mathbf{n}_k represents the complex Gaussian noise vector with zero mean and $\sigma^2 \mathbf{I}_{N_r}$.

By introducing the following matrices

$$\tilde{\mathbf{H}}_d = \overbrace{[\mathbf{h}_d, \mathbf{h}_d, \dots, \mathbf{h}_d]}^K \in \mathbb{C}^{N_r \times K}, \quad (7)$$

$$\tilde{\mathbf{H}}_2 = \overbrace{[\mathbf{H}_2, \mathbf{H}_2, \dots, \mathbf{H}_2]}^K \in \mathbb{C}^{N_r \times KN}, \quad (8)$$

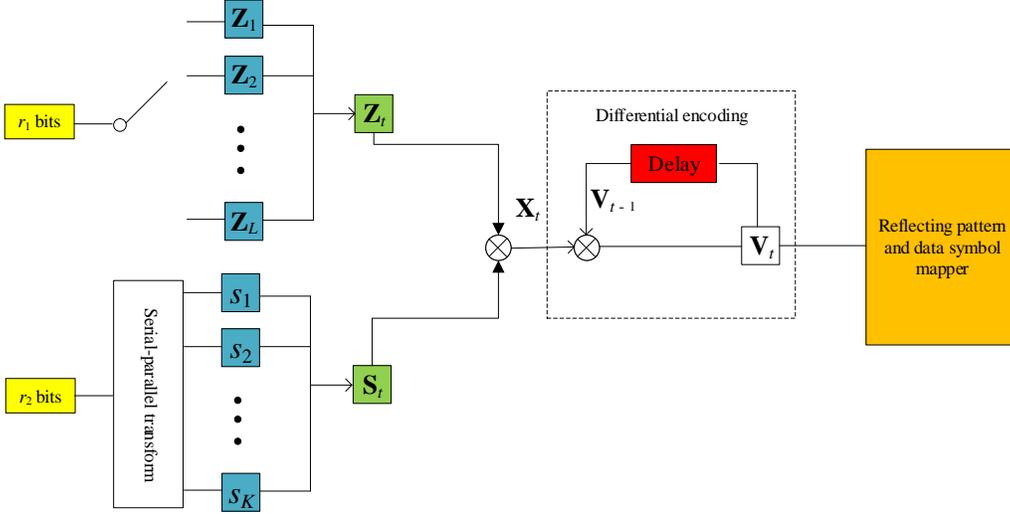


Fig. 2. DRM encoding process.

$$\mathbf{Q} = \begin{bmatrix} \Phi_1 & \mathbf{0} & \cdots & \mathbf{0} \\ \mathbf{0} & \Phi_2 & \cdots & \mathbf{0} \\ \vdots & \vdots & \ddots & \vdots \\ \mathbf{0} & \mathbf{0} & \cdots & \Phi_K \end{bmatrix} \in \mathbb{C}^{KN \times KN}, \quad (9)$$

$$\tilde{\mathbf{H}}_1 = \begin{bmatrix} \mathbf{h}_1 & \mathbf{0} & \cdots & \mathbf{0} \\ \mathbf{0} & \mathbf{h}_1 & \cdots & \mathbf{0} \\ \vdots & \vdots & \ddots & \vdots \\ \mathbf{0} & \mathbf{0} & \cdots & \mathbf{h}_1 \end{bmatrix} \in \mathbb{C}^{KN \times KN}, \quad (10)$$

we rewrite the expression of $\mathbf{y}_k^{(t)}$ as

$$\mathbf{y}_k^{(t)} = (\tilde{\mathbf{H}}_d + \tilde{\mathbf{H}}_2 \mathbf{Q} \tilde{\mathbf{H}}_1) \mathbf{v}_k^{(t)} + \mathbf{n}_k^{(t)}. \quad (11)$$

By setting $\mathbf{H} = \tilde{\mathbf{H}}_d + \tilde{\mathbf{H}}_2 \mathbf{Q} \tilde{\mathbf{H}}_1 \in \mathbb{C}^{N_r \times K}$, the transmission can be expressed as

$$\mathbf{y}_k^{(t)} = \mathbf{H} \mathbf{v}_k^{(t)} + \mathbf{n}_k^{(t)}, \quad (12)$$

where \mathbf{H} can be regarded as the equivalent channel matrix and $\mathbf{v}_k^{(t)}$ can be regarded as the equivalent transmit vector. Thus, during the t th block, the received signal matrix $\mathbf{Y}_t \in \mathbb{C}^{N_r \times K}$ can be expressed by

$$\mathbf{Y}_t = \mathbf{H} \mathbf{V}_t + \mathbf{N}_t, \quad (13)$$

where $\mathbf{Y}_t = [\mathbf{y}_1^{(t)}, \mathbf{y}_2^{(t)}, \dots, \mathbf{y}_K^{(t)}] \in \mathbb{C}^{N_r \times K}$ represents the received signal matrix, and $\mathbf{N}_t = [\mathbf{n}_1^{(t)}, \mathbf{n}_2^{(t)}, \dots, \mathbf{n}_K^{(t)}] \in \mathbb{C}^{N_r \times K}$ is the complex Gaussian noise matrix.

C. Detection Method

The advantage of the proposed DRM is that it can be detected without knowing any CSI at the transmitter, RIS or

receiver. The details are as follows. Substituting (2) into (13) yields

$$\begin{aligned} \mathbf{Y}_t &= \mathbf{H} \mathbf{V}_{t-1} \mathbf{X}_t + \mathbf{N}_t \\ &= \mathbf{Y}_{t-1} \mathbf{X}_t - \mathbf{N}_{t-1} \mathbf{X}_t + \mathbf{N}_t. \end{aligned} \quad (14)$$

Thus, the optimal maximum-likelihood (ML) detector can be derived as

$$\begin{aligned} \hat{\mathbf{X}}_t &= \arg \min_{\mathbf{X}_t \in \mathcal{X}} \|\mathbf{Y}_t - \mathbf{Y}_{t-1} \mathbf{X}_t\|_F^2 \\ &= \arg \min_{\mathbf{X}_t \in \mathcal{X}} \text{tr} \{(\mathbf{Y}_t - \mathbf{Y}_{t-1} \mathbf{X}_t)^H (\mathbf{Y}_t - \mathbf{Y}_{t-1} \mathbf{X}_t)\} \\ &= \arg \max_{\mathbf{X}_t \in \mathcal{X}} \Re \{ \text{tr}(\mathbf{Y}_t^H \mathbf{Y}_{t-1} \mathbf{X}_t) \}, \end{aligned} \quad (15)$$

where \mathcal{X} is the set of all legitimate \mathbf{X}_t and $|\mathcal{X}| = 2^r$. Then, information bits can be decoded from $\hat{\mathbf{X}}_t$ by according to the mapping rule given in Section III-A.

D. Transmission Rate and Complexity Analysis

Using DRM, the transmission rate can be written as

$$R = \frac{r}{K} = \frac{\lfloor \log_2 K! \rfloor + K \log_2 M}{K}, \quad (16)$$

in bits per channel use (bpcu). Based on the Stirling formula [18], $K! \approx \sqrt{2\pi K} (K/e)^K$, the transmission rate can be expressed as

$$R \approx \log_2 M + \lfloor \log_2 \sqrt{2\pi K} + K \log_2(K/e) \rfloor / K. \quad (17)$$

It can be checked that that the transmission rate increases as K increases, but the increase rate is not fast. The detection computational complexity can be analyzed to be

$$C_1 = 2^r (K^2 N_r + K^3) \text{ (multiplications)}, \quad (18)$$

since it needs to compute $\mathbf{Y}_t^H \mathbf{Y}_{t-1} \mathbf{X}_t$ by 2^r times and each computation requires $K^2 N_r + K^3$ multiplications. Recall that

$r = \lfloor \log_2 K! \rfloor + K \log_2 M$, the computational complexity can be expressed as

$$C_1 = (2^{\lfloor \log_2 K! \rfloor} + M^K)(K^2 N_r + K^3) \text{ (multiplications)}. \quad (19)$$

Based on the expression of C_1 , we find that the detection complexity increases much greatly as K increases.

E. Reflecting Pattern Selection

Above analysis indicates a good choice is to choose a small number of reflecting patterns for DRM transmission, which can enjoy low-complexity detection. Then, how to perform reflecting pattern selection in a finite set for transmission arouses our interest. Since CSI is neither known by the transceivers, nor by the RIS, we propose an optimization criterion to maximize the minimum mutual Euclidean distances, which is defined as

$$d_{\min} = \min_{\Phi_i, \Phi_{i'} \in \Psi, s_i, s_{i'} \in \mathcal{S}_M} \|\Phi_i s_i - \Phi_{i'} s_{i'}\|_2. \quad (20)$$

The rationale behind the optimization criterion is that that the information are jointly carried by the activation orders of the reflection pattern as well as the signal phases. In this letter, we adopt the stepwise depletion algorithm proposed in [3]. The detailed procedure of the stepwise depletion algorithm can be found in [3] and we omit it for brevity.

IV. SIMULATIONS

To show the performance of the proposed DRM scheme in (N_r, N, r) RIS-based SIMO communication systems, we first compare DRM with non-differential reflecting modulation (NDRM) using \mathbf{X}_t as transmission matrix with coherent detection. In NDRM, the received signal matrix $\hat{\mathbf{Y}}_t = \mathbf{H}\mathbf{X}_t + \mathbf{N}_t$, we adopt perfect $\mathbf{H} = \hat{\mathbf{H}}_d + \hat{\mathbf{H}}_2 \mathbf{Q} \hat{\mathbf{H}}_1$ (i.e., perfect \mathbf{H}_2 , \mathbf{h}_d and \mathbf{h}_1) for detection. The reason for choosing the comparison is that the transmission rate and detection complexity of both schemes are the same. In the comparison, we assume RIS is with $N = 4$ units with each unit being 1-bit encoded. That is, $(\Phi_k)_{nn} \in \{-1, 1\}$. Based on the assumption, we have $2^N = 16$ legitimate reflecting patterns for DRM and NDRM transmission. We chose two (i.e., $K = 2$) of them using the stepwise depletion algorithm [3]. The simulation results are depicted in Fig. 3. As the figure shows, DRM is less comparable to NDRM by 3 – 5 dB with the system setups as $N_r = 3$, and BPSK ($r = 3$) or QPSK ($r = 5$).

It is worth noting that DRM can work without CSI while NDRM have to spend much resources on channel estimation. Due to the fact that perfect CSI is typically not available, we further compare DRM and NDRM with imperfect CSI, where the imperfect CSI can be expressed by $\mathbf{H}_2^{\text{im}} = \mathbf{H}_2 + \mathbf{H}_2^e$, $\mathbf{h}_d^{\text{im}} = \mathbf{h}_d + \mathbf{h}_d^e$ and $\mathbf{h}_1^{\text{im}} = \mathbf{h}_1 + \mathbf{h}_1^e$. In the error model, \mathbf{H}_2^e , \mathbf{h}_d^e and \mathbf{h}_1^e represent the error terms, with each elements in the matrix or vectors following a complex Gaussian distribution with zero mean and covariance $\sigma_e^2 = \eta \sigma^2$ [3], where η is the positive proportional coefficient that is related to the number of pilots, the power and the employed algorithms for channel estimation. Simulation results are illustrated in Fig. 4, which demonstrate that the performance of NDRM reduces much as

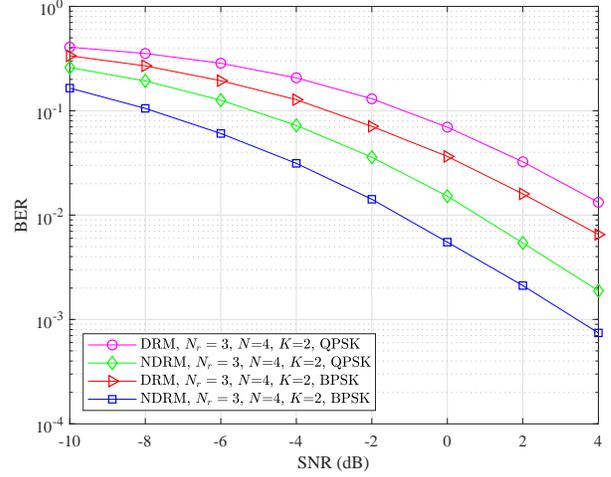


Fig. 3. Performance comparison in RIS-assisted (3,4,3) and (3,4,5) SIMO communication system with perfect CSI.

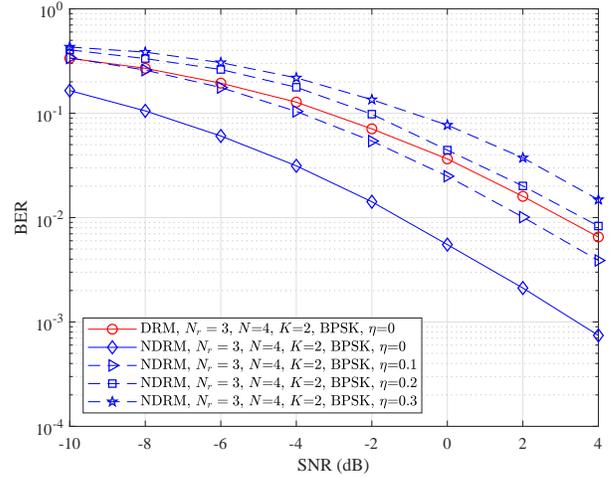


Fig. 4. Performance comparison in RIS-assisted (3,4,3) and (3,4,3) SIMO communication systems without perfect CSI.

CSI errors increase. The difference between the performance of DRM and that of NDRM becomes small when $\eta = 0.1$. When $\eta = 0.2$ and $\eta = 0.3$, DRM can outperform NDRM with coherent detection in the depicted SNR regime.

V. CONCLUSION

In this letter, a differential modulation scheme named as DRM was proposed for RIS-based communication systems. In DRM, the information bits are jointly encoded into the permutation order of the activated reflecting patterns and the phases of the transmit signals, which can be detected without knowing the channel state information. Simulation results showed that the difference between the proposed DRM and NDRM with coherent detection is acceptable, especially when there are inevitable channel estimation errors.

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