ICASSP 2023 - 2023 IEEE International Conference on Acoustics, Speech and Signal Processing (ICASSP) | 978-1-7281-6327-7/23/$31.00 ©2023 IEEE | DOI: 10.1109/ICASSP49357.2023.10096639

# HYBRID RIS-ASSISTED INTERFERENCE MITIGATION FOR SPECTRUM SHARING

*Fangzhou Wang and A. Lee Swindlehurst*

Center for Pervasive Communications and Computing, University of California Irvine

## ABSTRACT

This paper explores reconfigurable intelligent surfaces (RIS) for mitigating cross-system interference in spectrum sharing applications. Unlike conventional reflect-only RIS that can only adjust the phase of the incoming signal, a hybrid RIS is considered that can configure the phase and modulus of the impinging signal by absorbing part of the signal energy. We investigate two spectrum sharing scenarios: (1) Spectral coexistence of radar and communication systems, where a convex optimization problem is formulated to minimize the Frobenius norm of the channel matrix from the communication base station to the radar receiver, and (2) Spectrum sharing in device-to-device (D2D) communications, where a max-min scheme that optimizes the worst-case signal-to-interference-plus-noise ratio (SINR) among the D2D links is formulated, and then solved through fractional programming. Numerical results show that with a sufficient number of elements, the hybrid RIS can in many cases completely eliminate the interference, unlike a conventional non-absorptive RIS.

***Index Terms*—** Reconfigurable intelligent surfaces, interference mitigation, spectral coexistence, device-to-device communication.

## 1. INTRODUCTION

Reconfigurable Intelligent Surfaces (RIS), which can configure the wireless environment in a favorable manner by properly tuning the phase shifts of dozens of low-cost passive elements, has attracted significant attention within both the radar and wireless communication communities [1,2]. Some recent works on RIS include joint active and passive beamforming design [3], channel estimation [4], reflection modulation analysis [5], compressive sensing and deep-learning-based solutions [6], along with many other interesting application areas. In the radar community, RIS are employed to achieve non-line of sight surveillance [7], enhance the system reliability for target detection [8], improve parameter estimation performance [9], and suppress interference from clutter [10]. In addition, RIS were recently shown to play an important role for spectrum sharing between radar and communication systems [11]. RIS have been proposed to assist dual function radar communication (DFRC) transmission design to provide additional degree of freedoms (DoFs) to improve its performance [12–15]. In addition, the work in [16] and [17] focused on RIS-assisted interference mitigation for radar and communication system co-existence on shared spectrum.

In this paper, we consider the interference suppression problem for spectrum sharing by deploying a hybrid RIS. Unlike conventional RIS, which can only adjust the phase shift of the reflecting elements, or active RIS, which can also amplify the reflected signals using additional power consumption, here we propose to employ a hybrid RIS, which can adjust not only the phase but also the modulus of the reflecting elements by absorbing part of the signal energy. Note that the hardware complexity required by the hybrid RIS does not dramatically exceed that of the conventional RIS unless we want to leverage the absorbed signal energy for additional applications [18, 19], e.g., conduct local channel estimates or localization at the RIS. On the other hand, since hybrid RIS do not amplify the reflected signal, there is no additional noise introduced by the active components, as would be the case with an amplifier. Thus, the hybrid RIS serves to provide a trade-off between its active and passive counterparts in term of hardware complexity, power consumption, DoFs (e.g., adjustable modulus of the reflected signal), and the beamforming introduced by the RIS elements.

In this paper, the hybrid RIS assisted interference suppression problem is investigated under two spectrum sharing scenarios. The first explores the coexistence of both active and passive users on shared spectrum, i.e., the spectral coexistence of radar and communication systems, where the interference mitigation problem is formulated as minimizing the Frobenius norm of the channel matrix between a communication base station and a radar receiver (including both the direct channel and the indirect channel relayed by the hybrid RIS). The resulting optimization problem is proved to be a convex problem and can be efficiently solved by standard numerical solvers. The other examines the spectral coexistence of multiple device-to-device (D2D) users in the absence of a base station controller. A max-min criteria that optimizes the worst-case signal-to-interference-plus-noise ratio (SINR) among the D2D links is developed. This results in a non-convex constrained optimization problem, which is solved by an alternating Dinkelbach algorithm along with the semi-definite relaxation (SDR) technique. Numerical results

This work was supported by the National Science Foundation under grant CNS-2107182 and ECCS-2030029.

show that the proposed hybrid RIS can completely mitigate the interference by forcing all the interference elements of the channel matrix to zeros for both spectrum sharing scenarios, which cannot be achieved by only shifting the phase of the RIS elements as with conventional RIS.

## 2. COEXISTENCE OF RADAR AND COMMUNICATION SYSTEMS

In this section, we consider the interference suppression problem for the coexistence of radar and communication systems on shared spectrum with a deployed RIS.

### 2.1. Problem Formulation

Consider the coexistence of a communication base station (BS) with $M$ antennas and a passive radar RX with $N$ antennas using the same frequency band at the same time and/or at the same location. A hybrid RIS with $K$ reflecting elements is deployed near the radar RX to mitigate the interference from the BS to the radar RX. Unlike conventional RIS, which can only adjust the phase of the reflecting elements, a hybrid RIS can also adjust its modulus by absorbing part of the energy from the impinging signal.

Let $\alpha_k$ and $\rho_k \in [0\,1]$ denote the phase shift, and respectively, the modulus coefficient, associated with the $k$-th RIS element. Then, the diagonal matrix accounting for the RIS response can be expressed as $\boldsymbol{\Theta} \triangleq \text{diag}(\rho_1 e^{\jmath\alpha_1}, \cdots, \rho_K e^{\jmath\alpha_K})$. The channel between the BS and radar receiver can be written as $\mathbf{H}\boldsymbol{\Theta}\mathbf{G} + \mathbf{D}$, where $\mathbf{D} \in \mathbb{C}^{N\times M}$, $\mathbf{G} \in \mathbb{C}^{K\times M}$, and $\mathbf{H} \in \mathbb{C}^{N\times K}$ respectively denote the channel matrices between the BS and radar RX, the BS and the RIS, and the RIS and the radar RX. The problem of interest is to mitigate the interference channel from the BS to the radar RX by jointly designing the phase shift $\alpha_k$ and the absorbing modulus coefficient $\rho_k$ of the RIS. Specifically, the interference suppression problem can be formulated as

$$\min_{\{\rho_k,\alpha_k\}} \quad \left\|\mathbf{H}\boldsymbol{\Theta}\mathbf{G} + \mathbf{D}\right\|_F \tag{1a}$$

$$\text{s.t. } \boldsymbol{\Theta} = \text{diag}(\rho_1 e^{\jmath\alpha_1}, \cdots, \rho_K e^{\jmath\alpha_K}), \tag{1b}$$

$$0 \leq \rho_k \leq 1, \ \forall k. \tag{1c}$$

The extra DoF provided by $\rho_k$ enables the interference channel in many cases to be completely eliminated (i.e., $\mathbf{H}\boldsymbol{\Theta}\mathbf{G} + \mathbf{D} \to 0$) with a sufficiently large number of RIS elements. This is not possible by only adjusting the phase of the RIS elements as in conventional RIS.

### 2.2. Proposed Solution

While the interference mitigation problem in (1) is a nonlinear constrained optimization problem whose solution cannot be obtained directly, it can be reformulated to be convex and thus a global optimum can be found using standard techniques. In the following, we simplify the problem via mathematical manipulation. Specifically, by defining $\boldsymbol{\theta} \triangleq [\rho_1 e^{\jmath\alpha_1}, \cdots, \rho_K e^{\jmath\alpha_K}]^T$, it is easy to verify that the two constraints in (1b) and (1c) are equivalent to the following constraint: $|\boldsymbol{\theta}(k)| \leq 1, \ \forall k$, where $\boldsymbol{\theta}(k)$ is the $k$-th element of $\boldsymbol{\theta}$, i.e., $\boldsymbol{\theta}(k) = \rho_k e^{\jmath\alpha_k}$. It can be seen that the phase shifting parameter $\alpha_k$ and the modulus coefficient $\rho_k$ are combined into a single to-be-designed parameter $\boldsymbol{\theta}(k)$. Then, the optimization problem in (1) can be equivalently expressed as

$$\min_{\boldsymbol{\theta}} \quad \left\|\mathbf{H}\text{diag}(\boldsymbol{\theta})\mathbf{G} + \mathbf{D}\right\|_F, \ \text{ s.t. } |\boldsymbol{\theta}(k)| \leq 1, \ \forall k. \tag{2}$$

In addition, we have the following transformation for the cascaded channel: $\mathbf{H}\text{diag}(\boldsymbol{\theta})\mathbf{G} = \sum_{k=1}^{K} \boldsymbol{\theta}(k)\mathbf{H}(:,k)\mathbf{G}(k,:)$, where $\mathbf{H}(:,k)$ is the $k$-th column of $\mathbf{H}$ and $\mathbf{G}(k,:)$ is the $k$-th row of $\mathbf{G}$.

The cost function in (2) can be rewritten as

$$\begin{aligned} &\left\|\mathbf{D} + \sum_{k=1}^{K} \boldsymbol{\theta}(k)\mathbf{H}(:,k)\mathbf{G}(k,:)\right\|_F \overset{(a)}{=} \left\|\mathbf{D} + \sum_{k=1}^{K} \boldsymbol{\theta}(k)\mathbf{Y}_k\right\|_F \\ &\overset{(b)}{=} \left\|\text{vec}(\mathbf{D}) + \sum_{k=1}^{K} \boldsymbol{\theta}(k)\text{vec}(\mathbf{Y}_k)\right\|_2 \overset{(c)}{=} \left\|\mathbf{d} + \mathbf{A}\boldsymbol{\theta}\right\|_2, \end{aligned} \tag{3}$$

where $\mathbf{d} = \text{vec}(\mathbf{D})$. Note that the equality in (a) is obtained by letting $\mathbf{Y}_k = \mathbf{H}(:,k)\mathbf{G}(k,:)$, the equality in (b) is achieved by transferring the matrices into their vector forms, and the equality in (c) is due to the matrix definition that $\mathbf{A}$ is a $MN \times K$ matrix whose $k$-th column is given by $\mathbf{A}(:,k) = \text{vec}(\mathbf{Y}_k)$. Thus, the original optimization problem in (1) can be further rewritten as

$$\min_{\boldsymbol{\theta}} \quad \left\|\mathbf{d} + \mathbf{A}\boldsymbol{\theta}\right\|_2, \ \text{ s.t. } |\boldsymbol{\theta}(k)| \leq 1, \ \forall k, \tag{4}$$

which is a convex problem that can be efficiently solved.

## 3. DEVICE-TO-DEVICE COMMUNICATIONS

In this section, we investigate spectrum sharing for D2D communications where the hybrid RIS is deployed to suppress the cross interference among D2D links.

### 3.1. Problem Formulation

Device-to-device communications can increase network spectral efficiency, reduce transmission delay, and alleviate traffic congestion for the cellular infrastructure by enabling users in close proximity to transmit signals to each other directly without using the BS as a relay. In this section, we consider the cross-interference management problem among multiple single-antenna D2D communication links over shared spectrum with a deployed hybrid RIS, where $L$ D2D pairs coexist with each other in the same frequency band. Similar to the coexistence of radar and communication systems in Section 2, a hybrid RIS with reflecting response $\mathbf{\Theta}$ is deployed to mitigate the cross interference among the D2D links.

Let $\widetilde{\mathbf{D}} \in \mathbb{C}^{L\times L}$, $\widetilde{\mathbf{G}} \in \mathbb{C}^{K\times L}$, and $\widetilde{\mathbf{H}} \in \mathbb{C}^{L\times K}$ respectively denote the channel matrices between the transmitters and receivers, the transmitters and the RIS, and the RIS and the receivers, so that the overall channel between the transmitters and receivers of the D2D links can be expressed as $\widetilde{\mathbf{H}}\mathbf{\Theta}\widetilde{\mathbf{G}}+\widetilde{\mathbf{D}}$. Unlike the case in Section 2, to completely eliminate the interference channel $\mathbf{H\Theta G}+\mathbf{D}$, here we need to eliminate (or at least substantially reduce) the off-diagonal elements of the channel matrix $\widetilde{\mathbf{H}}\mathbf{\Theta}\widetilde{\mathbf{G}}+\widetilde{\mathbf{D}}$ since they represent the cross interference among the D2D links. To guarantee the performance of all D2D communication links, the cross-interference mitigation problem is formulated as maximizing the minimum SINR of the $L$ links by designing the parameters of the hybrid RIS, i.e.,

$$\max_{\{\rho_k,\alpha_k\}} \min_{\ell=1,\cdots,L} \frac{f_{\ell,\ell}(\rho_k,\alpha_k)}{\sum_{\bar{\ell}=1,\bar{\ell}\neq\ell}^{L} f_{\ell,\bar{\ell}}(\rho_k,\alpha_k)+\sigma^2} \tag{5a}$$

$$\text{s.t. (1b), (1c),} \tag{5b}$$

where $f_{\ell,\bar{\ell}}(\rho_k,\alpha_k) = \left|\widetilde{\mathbf{H}}(\ell,:)\mathbf{\Theta}\widetilde{\mathbf{G}}(:,\bar{\ell})+\widetilde{\mathbf{D}}(\ell,\bar{\ell})\right|^2 p_{\bar{\ell}}^2$, and $\sigma^2$ is the noise variance. Note that $p_\ell$ is the transmit power of the $\ell$-th link and $\widetilde{\mathbf{D}}(\ell,\bar{\ell})$ is the $(\ell,\bar{\ell})$-th element of $\widetilde{\mathbf{D}}$. The above problem is nonconvex and cannot be solved in closed-form. In the following, an iterative method is employed to solve it.

### 3.2. Proposed Solution

To solve (5), the definition of $\boldsymbol{\theta}(k) = \rho_k e^{\jmath\alpha_k}$ in Section 2 is employed to simplify the problem, in which case the constraint (5b) can be replaced by $|\boldsymbol{\theta}(k)| \leq 1$, $\forall k$ and $f_{\ell,\bar{\ell}}(\rho_k,\alpha_k)$ can be equivalently expressed as

$$\begin{aligned} f_{\ell,\bar{\ell}}(\rho_k,\alpha_k) &= \left|\widetilde{\mathbf{H}}(\ell,:)\mathbf{\Theta}\widetilde{\mathbf{G}}(:,\bar{\ell})+\widetilde{\mathbf{D}}(\ell,\bar{\ell})\right|^2 p_{\bar{\ell}}^2 \\ &= \left|\mathbf{h}_{\ell,\bar{\ell}}^H\boldsymbol{\theta}+\widetilde{\mathbf{D}}(\ell,\bar{\ell})\right|^2 p_{\bar{\ell}}^2 = \bar{\boldsymbol{\theta}}^H\mathbf{F}_{\ell,\bar{\ell}}\bar{\boldsymbol{\theta}}p_{\bar{\ell}}^2, \end{aligned} \tag{6}$$

where $\mathbf{h}_{\ell,\bar{\ell}}^*(k) = \widetilde{\mathbf{H}}(\ell,k)\widetilde{\mathbf{G}}(k,\bar{\ell})$, $\bar{\boldsymbol{\theta}} = [\boldsymbol{\theta}^T\ t]^T$, and

$$\mathbf{F}_{\ell,\bar{\ell}} = \begin{bmatrix} \mathbf{h}_{\ell,\bar{\ell}}\mathbf{h}_{\ell,\bar{\ell}}^H & \mathbf{h}_{\ell,\bar{\ell}}\widetilde{\mathbf{D}}_{\ell,\bar{\ell}} \\ \mathbf{h}_{\ell,\bar{\ell}}^H\widetilde{\mathbf{D}}_{\ell,\bar{\ell}}^* & |\widetilde{\mathbf{D}}_{\ell,\bar{\ell}}|^2 \end{bmatrix}. \tag{7}$$

The optimization problem in (5) can be rewritten as

$$\max_{\bar{\boldsymbol{\theta}}} \min_{\ell=1,\cdots,L} \frac{\bar{\boldsymbol{\theta}}^H\mathbf{F}_{\ell,\ell}\bar{\boldsymbol{\theta}}p_\ell^2}{\sum_{\bar{\ell}=1,\bar{\ell}\neq\ell}^{L}\bar{\boldsymbol{\theta}}^H\mathbf{F}_{\ell,\bar{\ell}}\bar{\boldsymbol{\theta}}p_{\bar{\ell}}^2+\sigma^2} \tag{8a}$$

$$\text{s.t. } |\bar{\boldsymbol{\theta}}(k)| \leq 1, k=1,\cdots,K, \text{ and } |\bar{\boldsymbol{\theta}}(K+1)| = 1. \tag{8b}$$

This is a fractional quadratically constrained quadratic programming (QCQP) problem and can be solved using SDR along with the Dinkelbach algorithm. In the following, we first use SDR to convert (8) to a fractional programming problem by dropping the rank-one constraint. Let $\bar{\mathbf{\Theta}} = \bar{\boldsymbol{\theta}}\bar{\boldsymbol{\theta}}^H$ to obtain the SDR form of (8) as

$$\max_{\bar{\mathbf{\Theta}}} \min_{\ell=1,\cdots,L} \frac{\mathrm{tr}(\bar{\mathbf{\Theta}}\mathbf{F}_{\ell,\ell})p_\ell^2}{\sum_{\bar{\ell}=1,\bar{\ell}\neq\ell}^{L}\mathrm{tr}(\bar{\mathbf{\Theta}}\mathbf{F}_{\ell,\bar{\ell}})p_{\bar{\ell}}^2+\sigma^2} \tag{9a}$$

$$\text{s.t. } |\bar{\mathbf{\Theta}}(k,k)| \leq 1, k=1,\cdots,K, \tag{9b}$$

$$|\bar{\mathbf{\Theta}}(K+1,K+1)| = 1, \tag{9c}$$

which is a fractional programming problem and can be solved by the Dinkelbach algorithm in polynomial time [20]. Specifically, by introducing a slack variable $\lambda$, problem (9) becomes (along with constraints (9b) and (9c))

$$\max_{\bar{\mathbf{\Theta}},\lambda} \min_{\ell=1,\cdots,L} \mathrm{tr}(\bar{\mathbf{\Theta}}\mathbf{F}_{\ell,\ell})p_\ell^2 - \lambda\Big(\sum_{\bar{\ell}=1,\bar{\ell}\neq\ell}^{L}\mathrm{tr}(\bar{\mathbf{\Theta}}\mathbf{F}_{\ell,\bar{\ell}})p_{\bar{\ell}}^2+\sigma^2\Big).$$

The above problem can be solved by alternatingly optimizing the cost function with respect to (w.r.t.) $\bar{\mathbf{\Theta}}$ and $\lambda$. In particular, by fixing $\lambda$ to the value obtained from the $i$-th iteration, $\lambda^{(i)}$, we can obtain $\bar{\mathbf{\Theta}}^{(i+1)}$ by solving (along with constraints (9b) and (9c))

$$\max_{\bar{\mathbf{\Theta}}} \min_{\ell=1,\cdots,L} \mathrm{tr}(\bar{\mathbf{\Theta}}\mathbf{F}_{\ell,\ell})p_\ell^2 - \lambda^{(i)}\Big(\sum_{\bar{\ell}=1,\bar{\ell}\neq\ell}^{L}\mathrm{tr}(\bar{\mathbf{\Theta}}\mathbf{F}_{\ell,\bar{\ell}})p_{\bar{\ell}}^2+\sigma^2\Big),$$

which is a convex problem. Next, we find $\lambda$ by fixing $\bar{\mathbf{\Theta}}$ to the value obtained from the latest update, $\bar{\mathbf{\Theta}}^{(i+1)}$, in which case $\lambda^{(i+1)}$ is solved in closed-form as $\lambda^{(i+1)} = \arg\min_{\ell,=1,\cdots,L} \frac{\mathrm{tr}(\bar{\mathbf{\Theta}}\mathbf{F}_{\ell,\ell})p_\ell^2}{\sum_{\bar{\ell}=1,\bar{\ell}\neq\ell}^{L}\mathrm{tr}(\bar{\mathbf{\Theta}}\mathbf{F}_{\ell,\bar{\ell}})p_{\bar{\ell}}^2+\sigma^2}$. The alternating process is repeated until convergence.

After solving (9) through the iterative Dinkelbach algorithm, we need to convert the optimum solution $\hat{\mathbf{\Theta}}$ to (9) into a feasible solution $\hat{\boldsymbol{\theta}}$ to (8). This can be achieved through a randomization approach [21]. Specifically, given the optimum $\hat{\mathbf{\Theta}}$, a set of Gaussian random vectors are generated, i.e., $\boldsymbol{\xi}_j \sim \mathcal{CN}(\mathbf{0},\hat{\mathbf{\Theta}})$, $j=1,\cdots,J$, where $J$ is the number of randomization trials. Since $\boldsymbol{\xi}_j$ are not always feasible for the modulus constraints in (8), we need to first normalize them through $\bar{\boldsymbol{\xi}}_j = \boldsymbol{\xi}_j/\boldsymbol{\xi}_j(K+1)$ to satisfy the constraint $|\bar{\boldsymbol{\theta}}(K+1)| = 1$. Then, to meet the constraint $|\bar{\boldsymbol{\theta}}(k)| \leq 1$, the feasible solution can be further recovered by $\hat{\boldsymbol{\xi}}_j(k) = e^{\jmath\angle(\bar{\boldsymbol{\xi}}_j(k))}$ if $|\bar{\boldsymbol{\xi}}_j(k)| > 1$. Finally, the rank-one solution can be obtained as $\hat{\boldsymbol{\theta}} = \arg\max_{\{\hat{\boldsymbol{\xi}}_j\}} \min_{\ell=1,\cdots,L} \frac{\hat{\boldsymbol{\xi}}_j^H\mathbf{F}_{\ell,\ell}\hat{\boldsymbol{\xi}}_j p_\ell^2}{\sum_{\bar{\ell}=1,\bar{\ell}\neq\ell}^{L}\hat{\boldsymbol{\xi}}_j^H\mathbf{F}_{\ell,\bar{\ell}}\hat{\boldsymbol{\xi}}_j p_{\bar{\ell}}^2+\sigma^2}$.

The computational complexity of the proposed algorithm mainly depends on the number of iterations of the Dinkelbach algorithm, i.e., $I$, and the number of SDR randomization trials $J$. Specifically, a convex problem is solved inside each iteration with a complexity of $\mathcal{O}((K+1)^{3.5})$ if an interior-point method is used, where $\mathcal{O}$ denotes the Landau notation. The computational complexity of the $J$ randomization trials is in the order of $\mathcal{O}(J(K+1)^2)$. Thus, the overall complexity of the proposed algorithm is $\mathcal{O}(I(K+1)^{3.5})+\mathcal{O}(IJ(K+1)^2)$.

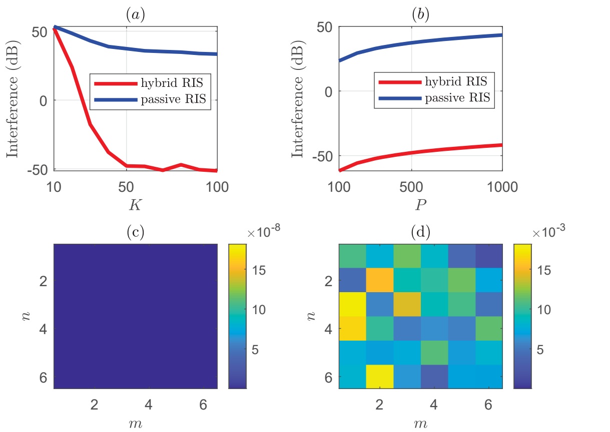


**Fig. 1**. Performance of RIS for interference suppression for radar/communication coexistence. (a) Interference versus number of RIS elements $K$; (b) Interference versus BS transmit power $P$; (c) Modulus of channel with hybrid RIS; (d) Modulus channel with non-absorptive RIS.

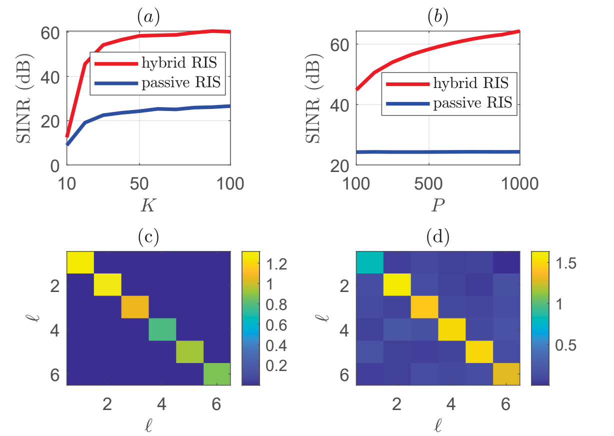


**Fig. 2**. Performance of RIS for interference suppression in D2D communications. (a) Minimum SINR versus $K$ when $p_\ell = P = 500$ and $L = 3$; (b) Minimum SINR versus $P$ when $K = 50$ and $L = 3$; (c) Modulus of channel with hybrid RIS when $K = 128$, $P = 500$ and $L = 6$; (d) Modulus of channel with non-absorptive RIS.

## 4. NUMERICAL RESULTS

In this section, simulation results are presented to evaluate the ability of hybrid RIS in interference mitigation. The performance of hybrid RIS is compared with that of the conventional non-absorptive RIS. In the simulation, Rayleigh fading channel models are utilized for $\mathbf{H}$, $\mathbf{G}$, $\mathbf{D}$, $\widetilde{\mathbf{H}}$, $\widetilde{\mathbf{G}}$, and $\widetilde{\mathbf{D}}$.

First, we consider the coexistence of a $M = 3$ antenna communication BS with a $N = 3$ antenna radar receiver. The interference from the BS to the radar receiver is given by $\|\mathbf{H}\boldsymbol{\Theta}\mathbf{G} + \mathbf{D}\|_F^2 P$ where $P$ is the BS transmit power. Fig. 1 (a) shows the interference energy versus the number of RIS elements $K$ when the BS transmit power is $P = 500$. It can be seen that the performance of both the hybrid and conventional RIS improves when the number of RIS elements $K$ increases. This is because a larger $K$ means more DoFs for the RIS to the handle the interference channel. In addition, the hybrid RIS outperforms the conventional RIS for all $K$, especially for larger RIS due to the availability of extra DoFs from the adaptivity of the RIS element modulus. The superiority of the hybrid over the conventional RIS is also shown in Fig. 1 (b), where the interference energy is plotted against the BS transmit power $P$ when $K = 50$. The interference energy for the hybrid RIS design remains at a negligible level (e.g., around $-50$ dB) even when the BS transmit power increases dramatically. This is because, in this case, when $K$ is large enough, the hybrid RIS is able to completely suppress the interference by forcing all the elements of the interference channel matrix to zero, i.e., $\mathbf{H}\boldsymbol{\Theta}\mathbf{G} + \mathbf{D} \to 0$, which cannot be achieved by using only phase shifts at the RIS. This is explicitly demonstrated by Fig. 1 (c) and (d) where the overall interference channel modulus is computed when $M = N = 6$ and $K = 128$. It can be observed that the hybrid RIS eliminates the interference channel by forcing all the channel gains to zero, which is not possible with a non-absorptive RIS.

Next, the performance of RIS in interference mitigation for the device-to-device communication scenario is shown in Fig. 2. A similar performance trend can be observed in terms of the output SINR of the communication links, i.e., larger output SINR for hybrid RIS than that for passive RIS which demonstrates their superior ability to achieve interference suppression. Unlike the case of radar and communication coexistence where it is desirable to zero out all the elements of the channel matrix between the BS and radar RX, here only the off-diagonal elements of the channel matrix $\widetilde{\mathbf{H}}\boldsymbol{\Theta}\widetilde{\mathbf{G}} + \widetilde{\mathbf{D}}$ are forced to zero since they represent the cross interference between D2D links, while the diagonal elements represent the gain of the desired communication links.

## 5. CONCLUSION

We examined the interference mitigation problem using hybrid RIS in two spectrum sharing scenarios, i.e., the coexistence of radar and communication systems, and device-to-device communications. Our main contributions include the problem formulations and the corresponding proposed solutions for each scenario. Our results indicate that hybrid RIS significantly outperform non-absorptive RIS for interference suppression scenarios. With a sufficient number of RIS elements, the hybrid RIS in many cases can force all the interference channel elements to zero by exploiting the extra DoFs (i.e., the modulus of the RIS elements) that are not available with conventional RIS.

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