Joint Antenna Selection and Passive Beamforming for an IRS Assisted Spectrum Sharing System

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***Abstract*—Combining intelligent reflecting surface (IRS), transmit antenna selection (AS), and spectrum sharing technologies reduces the hardware cost and enables efficient utilization of energy and the scarce spectrum. It improves the performance of an underlay spectrum sharing system by exploiting the diversity benefits of the multiple antennas and passive beamforming of IRS. We propose three different AS rules for a secondary transmitter (STx) that communicates to a secondary receiver with the help of an IRS while sharing spectrum with a primary system. The goal of the maximum signal power rule is to achieve constructive interference of the signals received at the secondary receiver through the direct link and the reflected link through IRS. On the other hand, the destructive interference rule aims at the destructive interference of the interference signals received at the primary receiver through the direct and the IRS reflected links from the STx. The interference satisfying rule aims to satisfy the interference constraint with equality. Our results show that the IRS assisted spectrum sharing system is different from the conventional spectrum sharing system without IRS and interference satisfying rule need not be the best choice always. Furthermore, we show that increase in the IRS elements improves the secondary performance significantly with only a marginal increase in the interference power at the primary.**

***Index Terms*—Intelligent reflecting surface, underlay, spectrum sharing, antenna selection, beamforming.**

I. Introduction

Intelligent reflecting surface (IRS) is a promising technology for sixth generation (6G) wireless systems that improves cost and energy efficiency [1]. It is made up of a large number of reflective elements whose reflection coefficients can be configured to modify the amplitude and phase of an incident signal. The reflection coefficients of these passive elements can be programmed to achieve passive beamforming, which improves the received signal power. Spectrum sharing enables efficient reuse of the limited spectrum. In underlay mode of spectrum sharing, the low priority secondary user transmits simultaneously in the same spectrum as the high priority primary user with constraints on the interference caused to the primary [2], [3].

In transmit antenna selection (AS), the transmitter selects one antenna from the set of antennas and connects it to the available radio frequency (RF) chain [3], [4]. It reduces the cost and complexity of the transmitter while retaining the spatial diversity of multiple antennas. It also enhances the performance of the secondary system, which is limited by interference constraint [2]. We aim to combine the benefits of AS, underlay spectrum sharing, and IRS to improve the efficiency in terms of the cost, spectrum, and energy.

Antenna Selection for an IRS Assisted System [1], [4], [5]: AS rules that jointly optimize the antenna indices at the transmitter and the reflection coefficients at the IRS to maximize the received power are developed in [1]. Furthermore, low-complexity AS rules that reduce computational complexity and the number of pilots required are also proposed. An AS rule is proposed for an IRS assisted single user wireless energy transfer system and its outage probability is derived in [5]. Joint AS at transmitter and passive beamforming at IRS using semidefinite relaxation technique to improve signal to noise ratio (SNR) at the receiver is developed in [4].

IRS Assisted Single Antenna Secondary System [6]–[9]: For an IRS assisted single-input-single-output (SISO) secondary system, joint binary power control and passive beamforming is studied to minimize the average symbol error probability (SEP) subject to an average interference constraint in [6]. The outage probability, ergodic capacity, and average BER analysis is done for an IRS assisted SISO secondary system without direct link and considering the interference from the primary transmissions to be Gaussian in [7]. Again, for a SISO secondary system that employs continuous power control, the joint optimization of the power at the secondary transmitter (STx) and passive beamforming at the IRS is done to maximize the achievable rate subject to the primary signal to interference plus noise ratio constraint [8], [9].

IRS Assisted Multiple Antenna Secondary System [10]–[13]: Joint transmit beamforming at the STx and passive beamforing at the IRS to maximize the secondary receiver's (SRx) rate is studied for a multiple-input-single-output (MISO) secondary system in [10]. It has been extended to multi-user system in [11]. A robust beamforming method is proposed in [12] to minimize the transmit power of an IRS assisted MISO spectrum sharing system. Penalty dual decomposition based gradient projection algorithm is proposed in [13] to maximize the rate at the SRx subject to the peak interference power constraint and unit modulus constraint.

A. Focus and Contributions

We focus on an IRS assisted underlay spectrum sharing system subject to the instantaneous interference constraint. In it, the STx equipped with multiple antennas performs

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AS to communicate to the SRx. It employs binary power control and transmits either with a fixed power or zero power. Our objective is to improve the secondary received signal strength while ensuring the instantaneous interference power seen at the primary receiver (PRx) is below a threshold. The novel aspects of our model are as follows: Firstly, we note that the joint AS and passive beamforming of an IRS assisted secondary system is not studied in the literature. Secondly, we consider a multiple antenna STx unlike the single antenna STx considered in [6]–[9]. Thirdly, we focus on binary power control, which reduces the complexity of the RF chain and enables design of an efficient power amplifier. Finally, we consider spatially correlated channel gains for the IRS links [1], which is different from independent channel gains considered in [5], [7].

Our specific contributions are as follows: For an IRS assisted underlay spectrum sharing system, we propose three novel AS rules that satisfy the interference constraint.

i) *Maximum signal power (MSP)* rule focuses on maximizing the received signal power at the SRx. For a given antenna at the STx, for each IRS element, it configures a reflection coefficient that aligns the direct and IRS reflected links from the STx to SRx. Then, the STx selects an antenna that yields the maximum received signal power among all the antennas that satisfy the instantaneous interference constraint.

ii) *Destructive interference (DI)* rule focuses on reducing the interference at the PRx. For each IRS element, it configures a reflection coefficient that aligns the direct STx-PRx link to be 180° out of phase to the STx-IRS-PRx reflected link. With these reflection coefficients, from the set of antennas that satisfy the interference constraint, the STx selects the antenna that maximizes the received signal power at the SRx.

iii) *Interference satisfying (IS)* rule focuses on satisfying the interference constraint with equality. Here, for each antenna at the STx, the reflection coefficients are obtained for solving an equation that equates the instantaneous interference to the interference threshold. Among the set of antennas for which solution is found, the STx selects the antenna that maximizes the received power at the SRx.

Both MSP and DI rules transmit with zero power if the set of antennas that satisfy the interference constraint is empty. Furthermore, the IS rule transmits with zero power if none of the antennas satisfy the interference constraint with equality.

Numerical Results: For a conventional underlay spectrum sharing system, the AS rule that meets the interference constraint with equality is shown to be the optimal [2]. However, we show that the IS rule need not be optimal for an IRS assisted spectrum sharing system. Our results also show that the IRS can improve the secondary performance $19.1\times$ while increasing the interference only by $1.2\times$ at the PRx as the number of IRS elements increases from 36 to 196.

Outline: Section II presents our system model. Our AS rules are discussed in Section III. Numerical results are presented in Section IV followed by conclusions in Section V.

Notations: Scalars are denoted by lower-case letters. Vectors and matrices are denoted by boldface lower-case and capital letters, respectively. Set of all complex-valued vectors of size $m \times 1$ is denoted $\mathbb{C}^{m\times 1}$ and $j = \sqrt{-1}$. A vector $\mathbf{x}$ with n^{th} element as x_n is denoted by $\mathbf{x} = [x_n]$ and $\mathbf{x}^T$ denotes its transpose. For a complex number a, $|a|$ and $\angle a$ denote its absolute value and angle, respectively.

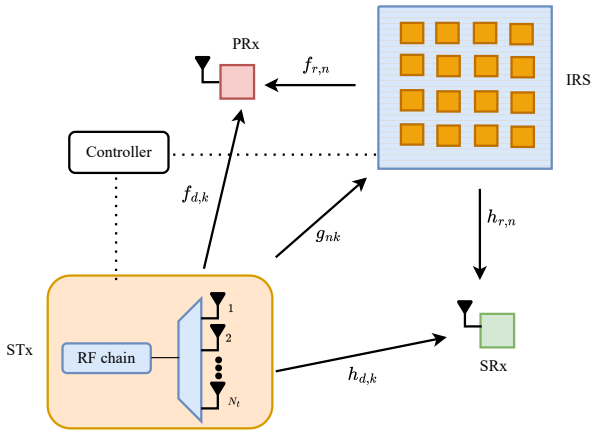


Fig. 1. System model consists of an IRS and STx equipped with N_t antennas and one RF chain communicates with SRx while sharing spectrum with the primary system.

II. System Model

In our system model, which is shown in Figure 1, the secondary system consists of an STx equipped with N_t antennas and one RF chain. It performs AS and communicates to a single antenna SRx with the help of an IRS equipped with N reflective elements. It shares spectrum with a primary system in underlay mode in which both primary and secondary can transmit simultaneously. The primary transmitter (PTx) and PRx are equipped with single antenna. The STx dynamically selects one antenna from the set $\{1, 2, \ldots, N_t\}$ and connects it to the available RF chain. The IRS controller receives the reflection coefficients from the STx through a control link. We denote the complex channel gains of the STx-SRx and STx-PRx direct links from the k^{th} transmit antenna of the STx by $h_{d,k}$ and $f_{d,k}$, respectively. Let g_{nk} denote the complex baseband channel gain from k^{th} STx antenna to the n^{th} IRS element and $\mathbf{g}_k = [g_{nk}] \in \mathbb{C}^{N\times 1}$ denote the channel gain vector from the k^{th} STx antenna to the IRS. Let $\mathbf{h}_r = [h_{r,n}] \in \mathbb{C}^{N\times 1}$ and $\mathbf{f}_r = [f_{r,n}] \in \mathbb{C}^{N\times 1}$ denote the complex baseband channel gain vectors from the IRS to the SRx and to the PRx, respectively.

Channel Model: The STx-SRx and STx-PRx direct links have only non line-of-sight (LoS) path. Therefore, we consider Rayleigh fading for the STx-SRx and STx-PRx direct links, i.e., $h_{d,k} \sim \mathcal{CN}(0, \mu_h^d)$ and $f_{d,k} \sim \mathcal{CN}(0, \mu_f^d)$, where μ_h^d and μ_f^d denote the mean channel power gains of the STx-SRx and STx-PRx links, respectively. We assume that the channel gains of STx-SRx, STx-PRx, and STx-IRS links corresponding to different transmit antennas are independent and identically distributed (i.i.d.) [1], [5]. IRS is positioned such that it has a dominant LoS to the STx, SRx, and PRx. Therefore, the

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STx-IRS, IRS-SRx, and IRS-PRx links are modeled as Rician random variables. We consider spatial correlation between the channel gains of different IRS elements corresponding to a given antenna at the STx, SRx, or PRx [1]. We use the correlation model proposed in [14]. The elements of the spatial correlation matrix are given by

$$[\mathbf{C}]_{n,m} = \text{sinc}\left(\frac{2d_{nm}}{\lambda}\right), \text{for } n, m = \{1, 2, \ldots, N\}, \quad (1)$$

where $\text{sinc}(x) = \sin(\pi x)/(\pi x)$ and d_{nm} is the spacing between the IRS elements n and m.

Therefore, the channel gain vectors IRS-SRx, IRS-PRx, and from the k^{th} STx antenna to the IRS are modeled as

$$\mathbf{h}_r = \sqrt{\frac{K\mu_r^h}{K+1}}\mathbf{h}_{\text{LoS}} + \sqrt{\frac{\mu_r^h \mathbf{C}}{K+1}}\mathbf{h}_{\text{NLoS}}, \quad (2)$$

$$\mathbf{f}_r = \sqrt{\frac{K\mu_r^f}{K+1}}\mathbf{f}_{\text{LoS}} + \sqrt{\frac{\mu_r^f \mathbf{C}}{K+1}}\mathbf{f}_{\text{NLoS}}, \quad (3)$$

$$\mathbf{g}_k = \sqrt{\frac{K\mu_r^g}{K+1}}\mathbf{g}_{\text{LoS}} + \sqrt{\frac{\mu_r^g \mathbf{C}}{K+1}}\mathbf{g}_{\text{NLoS}}, \quad (4)$$

where $\mathbf{h}_{\text{LoS}}$, $\mathbf{f}_{\text{LoS}}$, and $\mathbf{g}_{\text{LoS}}$ denote the LoS components and $\mathbf{h}_{\text{NLoS}}$, $\mathbf{f}_{\text{NLoS}}$, and $\mathbf{g}_{\text{NLoS}}$ denote non-LoS components. Each element of $\mathbf{h}_{\text{NLoS}}$, $\mathbf{f}_{\text{NLoS}}$, and $\mathbf{g}_{\text{NLoS}}$ vectors are distributed as i.i.d. $\mathcal{CN}(0,1)$. Furthermore, μ_r^h, μ_r^f, and μ_r^g denote the mean power gains of IRS-SRx, IRS-PRx and STx-IRS channel gains, respectively, and K denotes the Rician factor.

Let $x_n = \beta_n e^{j\theta_n}$ denote the reflection coefficient of the n^{th} IRS element, where $\theta_n \in [0, 2\pi]$ and $\beta_n \in [0, 1]$ denote the phase shift induced and amplitude gain, respectively. Let $\mathbf{x} = [x_1, \ldots, x_N]^T$ denote the passive beamforming vector at the IRS. The STx employs binary power control, which means it transmits either with fixed power $P_{\max}$ or zero power to avoid excessive interference to the primary [6]. It selects an antenna $s \in \{1, 2, \ldots, N_t\}$ and $P_s \in \{0, P_{\max}\}$ to transmit the data symbol d. The received signal at the SRx is given by

$$y = \sqrt{P_s}\left(h_{d,s} + \sum_{n=1}^{N} x_n h_{r,n} g_{ns}\right) d + z, \quad (5)$$

where $\mathbb{E}\left[|d|^2\right] = 1$, $z = n + I$, where n is circular symmetric complex additive white Gaussian noise at the SRx and I is the interference signal from the PTx, which we assume to be Gaussian [2], [3], [7]. Hence, $z \sim \mathcal{CN}(0, \sigma^2)$. Similarly, the interference signal received at the PRx is given by

$$I_p = \sqrt{P_s}\left(f_{d,s} + \sum_{n=1}^{N} x_n f_{r,n} g_{ns}\right) d. \quad (6)$$

Instantaneous Interference Constraint: From (6), the interference power at the PRx is given by

$$P_s \left| f_{d,s} + \sum_{n=1}^{N} x_n f_{r,n} g_{ns} \right|^2. \quad (7)$$

The instantaneous interference constraint mandates it to be below the threshold I_{th}.

Channel State Information Assumptions [15]–[19]: We assume that the STx knows the channel gains from itself to the SRx, i.e., $h_{d,1}, h_{d,2}, \ldots, h_{d,N_t}$ and cascaded channel gains through the IRS, i.e., $h_{r,n}g_{nk}$, $\forall n, \forall k$. The STx estimates these channel gains from the pilot symbols sent by the SRx using a two step procedure as described in [18], [19]. We also assume that the STx knows the direct and cascaded link channel gains from itself to the PRx, i.e., $f_{d,1}, f_{d,2}, \ldots, f_{d,N_t}$ and $f_{r,n}g_{nk}$, $\forall n$, $\forall k$. The STx can obtain this information by periodically sensing the signals from the PRx [17]. It can also exploit the hidden power feedback loop technique [16].

A. Problem Statement

For $P_s \in \{0, P_{\max}\}$, we aim to develop AS rules that select the transmit antenna index s and the passive beamforming vector $\mathbf{x}$ to improve the secondary received signal power

$$P_s \left| h_{d,s} + \sum_{n=1}^{N} x_n h_{r,n} g_{ns} \right|^2, \quad (8)$$

subject to the instantaneous interference constraint

$$P_s \left| f_{d,s} + \sum_{n=1}^{N} x_n f_{r,n} g_{ns} \right|^2 \leq I_{\text{th}}. \quad (9)$$

III. Antenna Selection With IRS

In this section, we will develop the MSP rule, which focuses on maximizing the received power, DI rule, which focuses on reducing interference at the PRx, and IS rule, which aims to meet the interference constraint with equality. For all the AS rules, we choose $\beta_n = 1$, $\forall n$ [8].

A. Maximum Signal Power Rule

This rule selects the transmit antenna, which yields the maximum signal power at the SRx and satisfies the interference constraint at the PRx, by configuring the reflection coefficient that ensures the signal received in the reflected link STx-IRS-SRx through each IRS element is in phase with the signal received in the STx-SRx direct link [1].

For an antenna $k \in \{1, 2, \ldots, N_t\}$ and n^{th} IRS element, we consider a reflection coefficient that aligns the direct signal and the reflected signal received at the SRx due to transmissions from the STx. It is given by

$$x_n^M = e^{j[\angle h_{d,k} - \angle(h_{r,n}g_{nk})]}. \quad (10)$$

For antenna k, this maximizes the signal power received at the SRx. Upon substituting this in (5) and (6), we note that the corresponding secondary received power at the SRx and the interference power received at the PRx are given by

$$P_{\max}\left(|h_{d,k}| + \sum_{n=1}^{N} |h_{r,n}g_{nk}|\right)^2 \quad (11)$$

and $P_{\max}\left|f_{d,k} + \sum_{n=1}^{N} x_n^M f_{r,n} g_{nk}\right|^2$, respectively. We now find out the set of antennas that satisfy the constraint. Let

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$\mathcal{F}$ denote the set of antennas that satisfy the interference constraint, i.e.,

$$\mathcal{F} = \left\{ k \middle| P_{\max} \left| f_{d,k} + \sum_{n=1}^{N} x_n^M f_{r,n} g_{nk} \right|^2 \leq I_{\text{th}} \right\}. \quad (12)$$

When $\mathcal{F}$ is not empty, we select the antenna $s \in \mathcal{F}$ that maximizes secondary received power, i.e.,

$$s = \arg\max_{k \in \mathcal{F}} \left\{ |h_{d,k}| + \sum_{n=1}^{N} |h_{r,n} g_{nk}| \right\}. \quad (13)$$

If $\mathcal{F}$ is empty, the STx transmits with zero power, i.e., $P_s = 0$. So the MSP rule ensures that the interference constraint is satisfied, by choosing $P_s = 0$, when none of the antennas satisfy the interference constraint ($\mathcal{F} = \{\}$) when x_n^M in (10) is configured.

B. Destructive Interference Rule

Unlike, the MSP rule, the DI rule focuses on reducing the interference at the PRx. To do that, for each antenna at the STx and IRS element, it computes a reflection coefficient that ensures the interference signal received at the PRx through the STx-IRS-PRx reflected link is 180° out of phase to the interference signal received in the STx-PRx direct link. For antenna $k \in \{1, 2, \ldots, N_t\}$, it is given by [8]

$$x_n^D = e^{j[\pi + \angle f_{d,k} - \angle (f_{r,n} g_{nk})]}. \quad (14)$$

Substituting this in (5) and (6), yields the signal power received at the SRx as $P_{\max} \left| h_{d,k} + \sum_{n=1}^{N} x_n^D h_{r,n} g_{nk} \right|^2$ and the destructive interference power I_{des} received at the PRx as

$$I_{\text{des}} = P_{\max} \left(|f_{d,k}| - \sum_{n=1}^{N} |f_{r,n} g_{nk}| \right)^2. \quad (15)$$

Let

$$\mathcal{D} = \left\{ k \middle| P_{\max} \left(|f_{d,k}| - \sum_{n=1}^{N} |f_{r,n} g_{nk}| \right)^2 \leq I_{\text{th}} \right\} \quad (16)$$

denote the set of antennas that satisfy the interference constraint upon configuring the reflection coefficient in (14). From the antennas in $\mathcal{D}$, we select the antenna s that maximizes received power at the SRx, i.e.,

$$s = \arg\max_{k \in \mathcal{D}} \left\{ \left| h_{d,k} - \sum_{n=1}^{N} e^{j[\angle f_{d,k} - \angle (f_{r,n} g_{nk})]} h_{r,n} g_{nk} \right|^2 \right\}. \quad (17)$$

The STx transmits with zero power, i.e., $P_s = 0$, if $\mathcal{D}$ is empty, which happens when none of the antennas satisfy the interference constraint when the reflection coefficient in (14) is configured.

C. Interference Satisfying Rule

Here, for each antenna k, we solve for the reflection coefficients that meet the interference constraint with equality, i.e.,

$$P_{\max} \left| f_{d,k} + \sum_{n=1}^{N} x_n f_{r,n} g_{nk} \right|^2 = I_{\text{th}}. \quad (18)$$

In order to understand the existence of the solution for the above equation, we consider the maximum interference and destructive interference possible at the PRx. The interference caused by the k^{th} STx antenna to the PRx is maximum when the x_n programmed aligns the STx-PRx direct link channel gain $f_{d,k}$ with the STx-IRS-PRx cascaded link channel gain $f_{r,n} g_{nk}$. This can be achieved by using the reflection coefficient $x_n = e^{j[\angle f_{d,k} - \angle (f_{r,n} g_{nk})]}$. Substituting this in (7), we get the corresponding maximum interference power $I_{\max}$ as

$$I_{\max} = P_{\max} \left(|f_{d,k}| + \sum_{n=1}^{N} |f_{r,n} g_{nk}| \right)^2. \quad (19)$$

Now the solution of (18) depends on the value of I_{th} relative to the $I_{\max}$ in (19) and I_{des} in (15) as follows:

1) $I_{\text{th}} > I_{\max}$: As $I_{\max}$ is the maximum interference power possible at the PRx, no solution exists for (18). Therefore, the IS rule configures the reflection coefficient based on (10), which maximizes the secondary received power while satisfying the interference constraint.

2) $I_{\text{des}} \leq I_{\text{th}} \leq I_{\max}$: Here, the interference threshold is less than the maximum interference power possible and greater than the destructive interference power. Therefore, a solution exists for (18) and we can obtain a passive beamforming vector $\mathbf{x}$ that satisfies the interference constraint with equality.

3) $I_{\text{th}} < I_{\text{des}}$: There are two possible cases under this condition [8]:

- (a) $|f_{d,k}| > \sum_{n=1}^{N} |f_{r,n} g_{nk}|$: Here, the strength of the STx-PRx direct link is larger than the effective cascaded STx-IRS-PRx link. Hence, no solution exists for (18) and the STx chooses $P_s = 0$.
- (b) $|f_{d,k}| < \sum_{n=1}^{N} |f_{r,n} g_{nk}|$: Here, a solution exists for (18) and we can obtain an $\mathbf{x}$ that meets the constraint.

Let $\mathcal{C}$ denote the set of antennas for which a solution exists for (18). Then, the index of the antenna, which maximizes the received signal power at the SRx is given by

$$s = \arg\max_{k \in \mathcal{C}} \left\{ \left| h_{d,k} + \sum_{n=1}^{N} x_n h_{r,n} g_{nk} \right| \right\}, \quad (20)$$

where x_n is obtained by solving (18).

IV. Numerical Results

In this section, we will study the impact of different system parameters on the performance of the proposed AS rules. Our simulation parameters are similar to that of [20]. The STx employs a uniform linear array with half wavelength spacing between the antennas and the IRS is made up of a uniform planar array with quarter wavelength sized elements.

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The correlation between IRS channel gains is computed using (1). The path-loss of different links is computed using $30 + 10\eta_p \log_{10}(d)$, where η_p denotes the path-loss exponent and d denotes the distance. We set $\eta_p = 3.6$ for the STx-SRx and STx-PRx direct links, which undergo Rayleigh fading, and $\eta_p = 2.2$ for the IRS links with strong LoS component. Furthermore, for the IRS links, we set the Rician factor $K = 10$. The IRS is placed at the origin and STx, SRx, and PRx are located at the distances of 70 m, 12 m, and 202 m, respectively, from it. Similarly, the distances of the SRx and PRx from the STx are equal to 72 m and 210 m, respectively. A carrier frequency of 2.4 GHz and $\sigma^2 = -104$ dBm are considered.

As no AS rules exist in the literature for IRS-assisted spectrum sharing, we benchmark the performance of the proposed AS rules with a *random phases* (RP) rule. Here, the IRS phase shifts are chosen randomly from a uniform distribution, i.e., $\theta_n \sim \mathcal{U}[0, 2\pi]$. For these random phases, an antenna that yields the maximum received power at the SRx is selected among the set of antennas that satisfy the interference constraint. It selects $P_s = 0$ if none of the antennas satisfy the constraint.

Figure 2 compares the performance of the proposed AS rules in terms of the average received signal power at the SRx normalized with noise power σ^2 as a function of the maximum transmit power $P_{\max}$ for different values of N_t. *Small* $P_{\max}$*:* The received signal power increases as $P_{\max}$ increases for all the AS rules. Here, the maximum interference is below the interference threshold I_{th} and IS rule is the same as that of the MSP rule. For that reason, the received power of the IS rule matches with the MSP rule. *Large* $P_{\max}$*:* Here, the received signal power of the MSP rule saturates as the STx transmits with zero power, i.e., $P_s = 0$, more often in order to meet the interference constraint. The saturation value increases as N_t increases. Furthermore, it is significantly higher than other rules. For $N_t = 8$ and $P_{\max} = 25$ dBm, the signal power for the MSP rule is 18 dB, 10 dB, and 13 dB higher than that of the DI, IS, and RP rules, respectively. A similar trend is observed for the RP rule with received power significantly lower than the MSP rule. For $P_{\max} = 40$ dBm, it is lower by 15 dB and 16 dB for $N_t = 2$ and $N = 8$, respectively. The DI rule exhibits a different trend, where the received signal power decreases and then increases. When the maximum interference exceeds I_{th}, the IS rule configures the reflection coefficients obtained by solving (18) and is different from the MSP rule. This leads to a transition in its performance at $P_{\max} = 15$ dBm. For this reason, the received power of the IS rule increases linearly with $P_{\max}$. The performance of the IS rule crosses the performance of the MSP rule for large $P_{\max}$. For $N_t = 2$, this happens at $P_{\max} = 40$ dBm.

Figure 3 plots the average interference power normalized by noise power σ^2 as a function of $P_{\max}$. For the IS rule, the interference power is independent of N_t for all $P_{\max}$. For small $P_{\max}$, it increases as $P_{\max}$ increases and saturates for large $P_{\max}$. The saturated value is equal to the interference threshold I_{th} as $\mathbf{x}$ is obtained by solving (18). *All AS Rules Other Than IS Rule:* The interference power increases as $P_{\max}$ increases initially and then decreases. This is because for small $P_{\max}$, the STx transmits with $P_s = P_{\max}$ more often and for large $P_{\max}$, it transmits with $P_s = 0$ more often to meet the interference constraint. For small $P_{\max}$, interference power is independent of N_t. For large $P_{\max}$, it increases as N_t increases as the STx transmits more often due to the availability of more antennas.

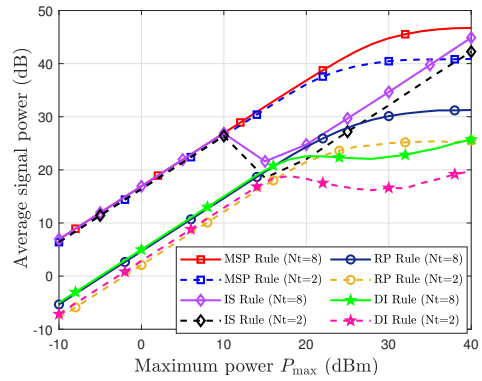


Fig. 2. Normalized average signal power at SRx as a function of $P_{\max}$ for different numbers of transmit antennas ($N = 64$ and $I_{\text{th}}/\sigma^2 = 3$ dB).

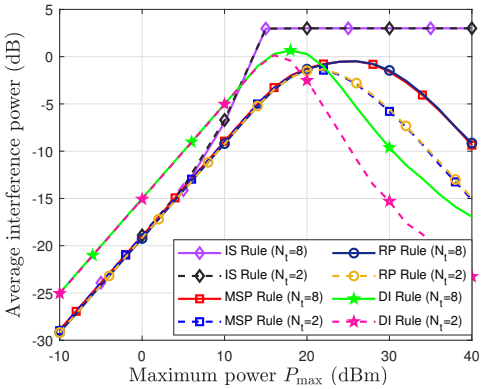


Fig. 3. Normalized average interference power at PRx as a function of $P_{\max}$ for different numbers of transmit antennas ($N = 64$ and $I_{\text{th}}/\sigma^2 = 3$ dB).

Figure 4 plots the average received signal power at the SRx as a function of I_{th}/σ^2 for different values of N. For the IS rule, the received signal power remains constant for smaller I_{th}. For larger I_{th} values, it matches with the MSP rule. This is because when $I_{\text{th}} > I_{\max}$, the IRS elements are configured with reflection coefficients same as the MSP rule. *All AS Rules Other Than IS Rule:* The signal power at the SRx increases as I_{th} increases and then saturates. This is because the probability of transmitting with $P_s = 0$ decreases as I_{th} increases, which relaxes the constraint, and reaches zero at a certain value of I_{th}. Beyond that value of I_{th}, the STx chooses $P_s = P_{\max}$ always and results in saturation.

Figure 5 plots the interference power at the PRx as a function of I_{th}/σ^2. For small values of I_{th}, the average interference power of the IS rule increases linearly with I_{th} as passive

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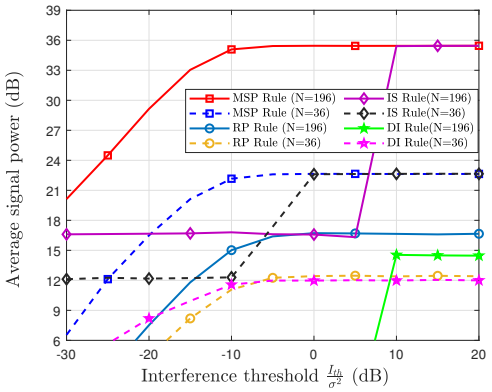


Fig. 4. Normalized average signal power at SRx as a function of I_{th}/σ^2 for different numbers of IRS elements ($P_{\max} = 10$ dBm and $N_t = 4$).

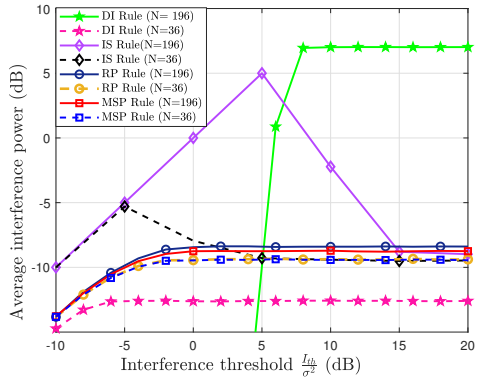


Fig. 5. Normalized average interference power at PRx as a function of I_{th}/σ^2 for different numbers of IRS elements ($P_{\max} = 10$ dBm and $N_t = 4$).

beamforming vector that satisfies the interference constraint with equality is chosen. For large I_{th}, it matches with the MSP rule. The average interference power of the MSP, DI, and RP rules increases as I_{th} increases and then saturates. We observe only a marginal increase in the interference power of the MSP and RP rules as N increases from 36 to 196. However, it increases significantly for the DI rule.

V. Conclusions

We proposed three new AS rules for an IRS assisted underlay spectrum sharing system that employs binary power control and is subject to the instantaneous interference constraint. The MSP rule that configures a reflection coefficient to align the direct link with the reflected link is shown to yield the best performance for small values of the transmit power. However, for large values of transmit power, the best performance was given by the IS rule, which ensured that the interference constraint was satisfied with equality and always transmitted with $P_{\max}$. We showed that the IS rule is not the best rule always unlike the conventional spectrum sharing systems without IRS. The increase in the average signal power as N_t or N increased is significantly higher for the MSP rule compared to the other AS rules. Furthermore, we saw that the increase in the number of IRS elements resulted in a significant increase in the received signal power with only a marginal increase in the interference power highlighting the benefit of the IRS for a spectrum sharing system.

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