Per-antenna Self-interference Cancellation Beamforming Design for Digital Phased Array

Ao Liu, *Student Member, IEEE*, Weixing Sheng, and Taneli Riihonen, *Senior Member, IEEE*

***Abstract*—Almost all current beamforming-based designs of self-interference cancellation (SIC) for digital phased array systems neglect the practical issue of the limited dynamic range of analog-to-digital converters (ADCs) and, thus, lack restriction for self-interference (SI) that is incident at each receive antenna. Other than beamforming, these methods usually require additional SIC techniques to provide a part of isolation, which brings higher system complexity. In this letter, a SIC method is proposed to overcome these issues. By developing a transmit SIC beamformer design that minimizes the power of SI on a per-antenna basis, we can obtain more isolation before the ADC in each receive channel and better prevent them all from being saturated. Then, the residual SI can be further suppressed by adaptive receive beamformers to achieve high total isolation performance. The proposed method reduces the need for other SIC techniques than beamforming technology in phased array systems and, thus, facilitates lower system complexity. Numerical results demonstrate the effectiveness of the proposed method.**

***Index Terms*—Simultaneous transmit and receive, in-band full-duplex, self-interference cancellation, digital beamforming.**

I. Introduction

SIMULTANEOUS transmit and receive (STAR) technology enables radio devices to perform without frequency division duplexing what the term suggests. It empowers unique system capabilities for various applications such as enhancing continuous-wave (CW) sensing performance for multi-mode radar systems [1]–[3], realizing in-band full-duplex (IBFD) communications [4]–[6], and co-designing emerging joint communication and sensing (JCAS) systems [7]–[9] with mutual benefits for both functions. Digital phased arrays with a transceiver behind each antenna are usually required to implement such applications with the STAR capability.

To implement STAR, the self-interference (SI) from a transmitting (Tx) subarray to a co-located receiving (Rx) subarray has to be sufficiently mitigated [10] — otherwise the coupled SI and noise will saturate the receivers and hamper the proper functioning of the system. Multiple SI cancellation (SIC) techniques have been proposed at the propagation domain [11]–[15], analog domain [16], and digital domain [17]. For SIC in multi-channel arrays, beamforming is the prevalent technology. Except for [18], all existing methods for SI cancellation (SIC) beamforming neglect, to the authors' knowledge, the practical effect of the SI signal on analog-to-digital converters (ADCs) with limited dynamic range [19]–[22]. Consequently, they lack constraints for the incident SI signal at each Rx antenna, which may cause the saturation of ADCs in adjacent receivers and the effective resolution of a weak signal-of-interest (SOI) will be heavily decreased. Especially, those methods then need extra SIC techniques in addition to beamforming to provide a part of isolation [20]–[22], which will lead to higher system complexity.

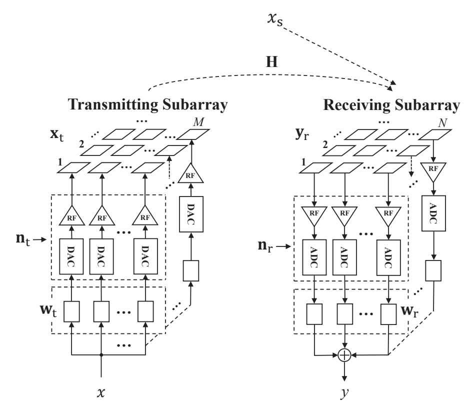

Fig. 1. Signal flow diagram of the proposed method design with M Tx antennas and N Rx antennas.

In this letter, a novel SIC method that exploits digital beamforming (DBF) technology is proposed to reduce more the SI that incidents at each Rx antenna with improved effective ADC resolution and achieve lower system complexity. First, a transmit beamformer (BF) design that minimizes the power of SI signal that is coupled to each Rx antenna is developed by converting the non-convex minimization problem into second-order cone programming (SOCP). This per-antenna design effectively mitigates the SI signal before the ADC of each Rx channel to better prevent all receivers from being saturated and, thus, higher total isolation before Rx subarray is also achieved. Then, an adaptive receive BF is proposed to further cancel the residual SI in Rx channels.

In Section IV, numerical results demonstrate the improved performance of the proposed method. With significant total isolation achieved, the proposed method reduces the need for any other SIC technique and, thus, renders lower system complexity. Meanwhile, it has the potential to be implemented in any digital phased array and extended to different application scenarios, as only beamforming technique is exploited.

II. System Model

Let us consider a digital phased array system, where the aperture is partitioned into a Tx subarray and an adjacent Rx

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A. Liu and W. Sheng are with the School of Electronic and Optical Engineering, Nanjing University of Science and Technology, 210094 Nanjing, China. (e-mail: liuao@njust.edu.cn; shengwx@njust.edu.cn).

T. Riihonen is with the Faculty of Information Technology and Communication Sciences, Tampere University, 33720 Tampere, Finland (e-mail: taneli.riihonen@tuni.fi).

subarray, as shown in Fig. 1. In this letter, we focus on this adjacent partitioning, as it offers better overall performance than other options, such as interleaved or random ones as shown in [20], [21]. We denote that there are M Tx antennas and N Rx antennas in the partitioned array.

When $x(k) \in \mathbb{C}$ is the time-varying signal to be transmitted with unit power $\mathbf{E}\left[|x(k)|^2\right] = 1$, the Tx signal $\mathbf{x}_\mathrm{t}(k) \in \mathbb{C}^M$ at the kth snapshot is

$$\mathbf{x}_\mathrm{t}(k) = \mathbf{w}_\mathrm{t} x(k) + \mathbf{n}_\mathrm{t}(k) \tag{1}$$

where $\mathbf{w}_\mathrm{t} \in \mathbb{C}^M$ is the transmit weight with total transmitted power $P_\mathrm{t} = \|\mathbf{w}_\mathrm{t}\|_2^2$ and $\|\cdot\|_2$ denotes the l_2-norm. In practical transmitter hardware, $\mathbf{n}_\mathrm{t}(k) \in \mathbb{C}^M$ represents zero-mean, additive white Gaussian noise (AWGN) due to limited dynamic range with covariance matrix $\mathbf{N}_\mathrm{t} = \mathbf{E}\left[\mathbf{n}_\mathrm{t}(k)\mathbf{n}_\mathrm{t}^H(k)\right] = \mathrm{diag}\left(\mathbf{E}[(\mathbf{w}_\mathrm{t}x(k))(\mathbf{w}_\mathrm{t}x(k))^H]\right)/\eta_\mathrm{t}$, where η_t is the signal-to-noise ratio (SNR) of each transmitter [23]. While $\mathrm{diag}(\cdot)$ denotes the diagonal matrix, $\mathbf{E}$ denotes statistical expectation and $(\cdot)^H$ denotes the conjugate transpose.

The Rx signal after receive weight is

$$y(k) = \mathbf{w}_\mathrm{r}^H(\mathbf{a}_\mathrm{r} x_\mathrm{s}(k) + \mathbf{H}\mathbf{x}_\mathrm{t}(k) + \mathbf{n}_\mathrm{r}(k)) \tag{2}$$

where $x_\mathrm{s}(k) \in \mathbb{C}$ is the incident SOI, $\mathbf{a}_\mathrm{r} \in \mathbb{C}^N$ is the corresponding Rx subarray manifold and $\mathbf{w}_\mathrm{r} \in \mathbb{C}^N$ is the receive weight with $\|\mathbf{w}_\mathrm{r}\|^2 = 1$. The SI channel matrix is represented by $\mathbf{H} \in \mathbb{C}^{N\times M}$. The SI channel estimation error is considered and the imperfect SI channel matrix can be expressed as $\hat{\mathbf{H}} = \mathbf{H} + \mathbf{H}_\mathrm{e}$ [24], [25], where $\mathbf{H}_\mathrm{e} \in \mathbb{C}^{N\times M}$ is the estimation error matrix and each entry is modeled as zero-mean AWGN with variance [25]

$$\varepsilon_{i,j}^2 = \epsilon^2 \mathbf{E}\left[|\mathbf{H}_{i,j}|^2\right] \tag{3}$$

for $i = 1, \ldots, N$ and $j = 1, \ldots, M$, where ϵ is the relative estimation error. In practical receiver hardware, $\mathbf{n}_\mathrm{r}(k) \in \mathbb{C}$ is zero-mean AWGN [23] with covariance matrix $\mathbf{N}_\mathrm{r} = \mathbf{E}\left[\mathbf{n}_\mathrm{r}(k)\mathbf{n}_\mathrm{r}^H(k)\right] = \mathrm{diag}\left(\mathbf{E}\left[\mathbf{y}_\mathrm{r}(k)\mathbf{y}_\mathrm{r}^H(k)\right]\right)/\eta_\mathrm{r} + \sigma_\mathrm{r}^2\mathbf{I}_N$, where $\mathbf{y}_\mathrm{r}(k) = \mathbf{a}_\mathrm{r}\mathbf{x}_\mathrm{s}(k) + \mathbf{H}\mathbf{x}_\mathrm{t}(k)$, $\mathbf{I}_N$ denotes the $N\times N$ identity matrix, and η_r is the SNR of each receiver.

III. Proposed Method

A. Transmit SIC Design

The objectives of our transmit SIC design are (i) to mitigate the coupled SI signal that is incident at each Rx antenna to protect the receivers' ADCs and (ii) to maintain high Tx gain in desired direction. Therefore, we aim to optimize the transmit beamformer that minimize the maximum SI signal power on all Rx antennas. When the maximum SI signal power is minimized, the SI power at each Rx antenna will be spontaneously minimized as expected. First, the SI power on ith Rx antenna can be expressed as $\mathbf{E}[|\mathbf{h}_i\mathbf{w}_\mathrm{t}x(k)|^2] = |\mathbf{h}_i\mathbf{w}_\mathrm{t}|^2$, where $\mathbf{h}_i \in \mathbb{C}^{1\times M}$ denotes the ith row of the SI channel matrix $\mathbf{H}$. Thus, an optimization problem can be formulated as

$$\min_{\mathbf{w}_\mathrm{t}} \quad \max_{1\le i\le N} |\mathbf{h}_i\mathbf{w}_\mathrm{t}|^2 \tag{4}$$

to reach the foregoing objective.

The problem can be further formulated as

$$\mathcal{P}_1: \begin{cases} \min\limits_{\mathbf{w}_\mathrm{t}} & \max\limits_{1\le i\le N} |q_i|^2 \\ s.t. & \frac{g(\theta,\phi)\mathbf{w}_\mathrm{t}^H\mathbf{a}_\mathrm{t}^H(\theta,\phi)\mathbf{a}_\mathrm{t}(\theta,\phi)\mathbf{w}_\mathrm{t}}{\mathbf{w}_\mathrm{t}^H\mathbf{w}_\mathrm{t}} \ge \gamma \end{cases} \tag{5}$$

where $q_i = \mathbf{h}_i\mathbf{w}_\mathrm{t} \in \mathbb{C}$. The constraint is added to guarantee the gain is not less γ, where $\mathbf{a}_\mathrm{t}(\theta,\phi)$ and $g(\theta,\phi)$ are the steering vector and embedded element gain towards (θ,ϕ), respectively.

$\mathcal{P}_1$ is hard to be directly solved, since the cost function in $\mathcal{P}_1$ is non-differentiable and the constraint is non-convex. However, we show that $\mathcal{P}_1$ can be further converted into a solvable SOCP, which finally is transformed as a per-antenna design that minimizes the SI power on each Rx antenna. Therefore, the complex-valued problem $\mathcal{P}_1$ is further transformed as a real-valued problem

$$\mathcal{P}_2: \begin{cases} \min\limits_{\mathbf{w}_\mathrm{t}} & u \\ s.t. & \Re\{q_i\}^2 + \Im\{q_i\}^2 \le u, i = 1, ..., N \\ & \mathbf{a}_\mathrm{t}'\mathbf{w}_\mathrm{t}' = \begin{bmatrix} c \\ 0 \end{bmatrix} \\ & {\mathbf{w}'}_\mathrm{t}^H\mathbf{w}_\mathrm{t}' \le \frac{1}{\gamma} \end{cases} \tag{6}$$

where

$$\mathbf{w}_\mathrm{t}' = \begin{bmatrix} \Re(\mathbf{w}_\mathrm{t}) \\ \Im(\mathbf{w}_\mathrm{t}) \end{bmatrix}, \quad \mathbf{a}_\mathrm{t}' = \begin{bmatrix} \Re(\mathbf{a}_\mathrm{t}) & -\Im(\mathbf{a}_\mathrm{t}) \\ \Im(\mathbf{a}_\mathrm{t}) & \Re(\mathbf{a}_\mathrm{t}) \end{bmatrix} \tag{7}$$

and $c = 1/\sqrt{g(\theta,\phi)}$ is a constant. The non-convex constraint in $\mathcal{P}_1$ for gain is converted into two convex formulations in $\mathcal{P}_2$. The parameter γ is adjustable for tradeoff between Tx gain and achieved SIC. Meanwhile, $\Re$ and $\Im$ denote the real and imaginary operators respectively, where ${\mathbf{w}_\mathrm{t}}' \in \mathbb{R}^{2M}$ and ${\mathbf{a}_\mathrm{t}}' \in \mathbb{R}^{2\times 2M}$. $\mathcal{P}_1$ is finally converted to the per-antenna scheme $\mathcal{P}_2$ of minimizing SI power on each Rx antenna, which is formulated to be less than an auxiliary variable $u \in \mathbb{R}^+$ as $\Re\{q_i\}^2 + \Im\{q_i\}^2 \le u, i = 1, ..., N$ and u will be minimized.

The problem $\mathcal{P}_2$ is a well-formulated SOCP problem and this per-antenna design aims to reduce SI signals on all Rx channels. Since SI signals incident at all Rx channels are effectively mitigated, the total SI power coupled to Rx subarray is also minimized. Compared with previously proposed methods, our design further regulates the SI signal on each Rx antenna to prevent all receivers from being saturated and achieve more isolation. Then, $\mathcal{P}_2$ can be well solved by MOSEK or CVX with numerical solution of transmit weight ${\mathbf{w}_\mathrm{t}}'$. It is worth noting that the proposed method can be further applied to wideband scenario, such as for wideband orthogonal frequency division multiplexing systems, as the proposed design can be parallel applied for each subcarrier thereof.

B. Receive SIC Design

In this letter, the EII metric [26], which is the ratio of effective isotropic radiated power (EIRP) to effective isotropic sensitivity (EIS), is adopted to comprehensively measure both the achieved total isolation and gains of a array system. $\mathrm{EIRP} = P_\mathrm{t}G_\mathrm{t}$, where G_t is the gain of Tx subarray; $\mathrm{EIS} =$

$P_{\mathrm{n}}/G_{\mathrm{r}}$, where P_{n} is the residual noise and interference power after receive BFs and G_{r} is the gain of Rx subarray.

After the transmit SIC, the EII can be formulated as

$$\mathrm{EII} = \mathrm{EIRP} \cdot \frac{G_{\mathrm{r}}}{P_{\mathrm{n}}} \tag{8}$$

where EIRP is constant as the transmit weight has been determined. To further achieve significant total isolation w.r.t. EII, it is explicit that we need to maximize (8), which is to minimize P_{n} and maximize G_{r}. According to (2), the residual SI and noise signals are composed of a SI signal due to the Tx signal and transmitter noise and receiver noise. The residual SI and noise covariance matrix is

$$\begin{aligned}\mathbf{C}_{\mathrm{r}} =& \mathbf{H}\mathbf{w}_{\mathrm{t}}\mathbf{w}_{\mathrm{t}}^{H}\mathbf{H}^{H} \\ &+ \eta_{\mathrm{r}}^{-1}\,\mathrm{diag}\left(\mathbf{H}\mathbf{w}_{\mathrm{t}}\mathbf{w}_{\mathrm{t}}^{H}\mathbf{H}^{H}\right) \\ &+ \eta_{\mathrm{t}}^{-1}\mathbf{H}\,\mathrm{diag}\left(\mathbf{w}_{\mathrm{t}}\mathbf{w}_{\mathrm{t}}^{H}\right)\mathbf{H}^{H} \\ &+ \eta_{\mathrm{r}}^{-1}\eta_{\mathrm{t}}^{-1}\,\mathrm{diag}\left[\mathbf{H}\,\mathrm{diag}\left(\mathbf{w}_{\mathrm{t}}\mathbf{w}_{\mathrm{t}}^{H}\right)\mathbf{H}^{H}\right] \\ &+ \sigma_{\mathrm{r}}^{2}\mathbf{I}_{N}\end{aligned} \tag{9}$$

Then, the residual SI and noise power can be formulated as

$$P_{\mathrm{n}} = \mathbf{w}_{\mathrm{r}}^{H}\mathbf{C}_{\mathrm{r}}\mathbf{w}_{\mathrm{r}} \tag{10}$$

The expression of EII can be formulated as

$$\mathrm{EII} = \mathrm{EIRP} \cdot g(\theta,\phi) \cdot \frac{\mathbf{w}_{\mathrm{r}}^{H}\mathbf{a}_{\mathrm{r}}(\theta,\phi)\mathbf{a}_{\mathrm{r}}^{H}(\theta,\phi)\mathbf{w}_{\mathrm{r}}}{\mathbf{w}_{\mathrm{r}}^{H}\mathbf{C}_{\mathrm{r}}\mathbf{w}_{\mathrm{r}}} \tag{11}$$

where $\|\mathbf{w}_{\mathrm{r}}\|_2^2 = 1$. Note that the minimum variance distortionless response (MVDR) criterion can be applied to efficiently solve the maximization problem in (11). Therefore, the closed-form solution of the adaptive receive weight is

$$\mathbf{w}_{\mathrm{r}} = \alpha\mathbf{C}_{\mathrm{r}}^{-1}\mathbf{a}_{\mathrm{r}} \tag{12}$$

where α is the arbitrary scale factor to satisfy $\|\mathbf{w}_{\mathrm{r}}\|_2^2 = 1$.

Finally, the computational complexity of the proposed method is about $\mathcal{O}(T_1M^{3.5} + N^3 + (N^2M + M^2))$, where $\mathcal{O}(T_1M^{3.5})$ is the worst-case complexity of solving SOCP transmit BF design with T_1 iterations. $\mathcal{O}(N^3)$ is the complexity of MVDR receive BF design and $\mathcal{O}(N^2M + M^2)$ is the complexity of constructing covariance matrix. Moreover, the indirect multipath model can also be applied to the proposed method without interfering with the problem formulations and solutions.

IV. Performance Analysis

A. Parameter Configuration

A 12×6 U-slot patch antenna array was modeled in Ansys HFSS. The array is half-wavelength spaced and operates at 28 GHz center frequency with narrow bandwidth. The array is partitioned into 6×6 Tx and Rx subarrays, as shown in Fig. 1. Mutual coupling $\mathbf{H}$ was estimated by Ansys HFSS. $\eta_{\mathrm{t}} = 50$ dB, $\eta_{\mathrm{r}} = 70$ dB. For uniform BFs, 10 W is set as the upper limit of transmitted power of each Tx channel, which renders 360 W as the upper limit of the total transmitted power P_{t} for the numerical results.

Due to the tradeoff feature, γ will be decided according to different system requirements but not a fixed optimal value.

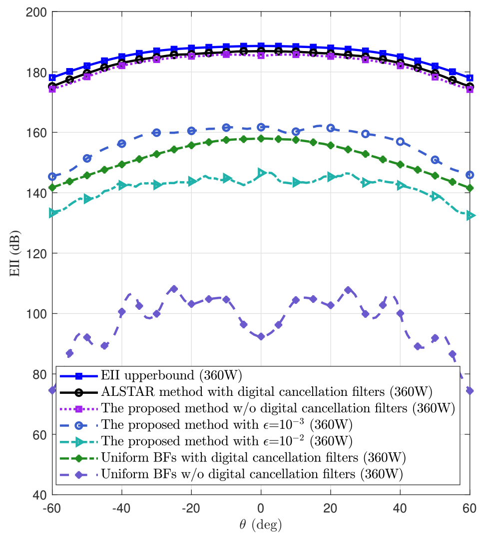

Fig. 2. EII plotted across θ scan angle for EII upper bound, uniform BFs without any SIC technique, uniform BFs with digital cancellation filters, ALSTAR method with digital cancellation filters and the proposed method without digital cancellation filters for various ϵ.

For the following numerical results, the γ is set as 2 dB lower than the gain of uniform BFs, which indicates about 2 dB Tx gain is traded off for SIC performance. The aforementioned exceptional method in [18] is designed for a hybrid beamforming architecture, which does not allow appropriate comparison with our method. Therefore, the ALSTAR in [21] is selected as the reference scheme, as their method also exploits DBF technique at both sides to mitigate SI in shared-aperture digital phased arrays and yields significant SIC performance. However, their method requires the additional digital cancellation filters with auxiliary Rx chains.

The upper bound of EII, which is

$$\mathrm{EII}_{\mathrm{ub}} = \frac{P_{\mathrm{t}}MPg^2(\theta,\phi)}{\sigma_{\mathrm{r}}^2} \tag{13}$$

is exploited to evaluate the total isolation performance. Compared to receiver operation without SI, the bit loss of ADC due to SI and noise in each Rx channel is exploited to evaluate the effectiveness of transmit BFs [27]. This parameter directly quantify the amount of dynamic range that is reserved by the SI and noise. Higher bit loss indicates more SI and noise are incident at the Rx channel and inferior SIC performance of transmit BFs. On the contrary, lower bit loss indicates higher isolation is achieved by transmit BFs and more dynamic range of ADC is preserved for weak desired signal.

B. Numerical Results

Several numerical results are presented to demonstrate the benefits of the proposed method. The achieved EII of the uniform BFs without SIC techniques, uniform BFs with digital cancellation filters, ALSTAR method with digital cancellation filters and the proposed method without digital cancellation filters are compared across the $\pm 60^\circ$ scan angles as show

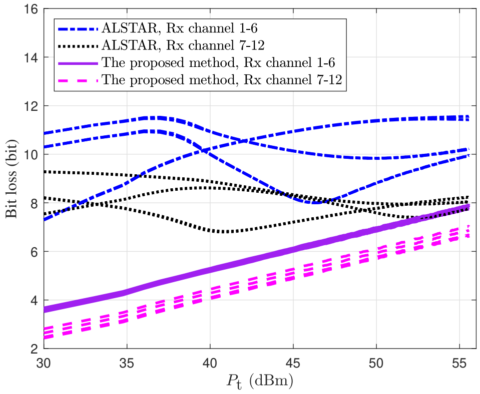

Fig. 3. Comparison for bit loss of ADCs in 12 Rx channels that adjacent to the Tx subarray between the proposed transmit BFs and ALSTAR transmit BFs from low transmitted power level to high transmitted power level, when the beam is steered at broadside.

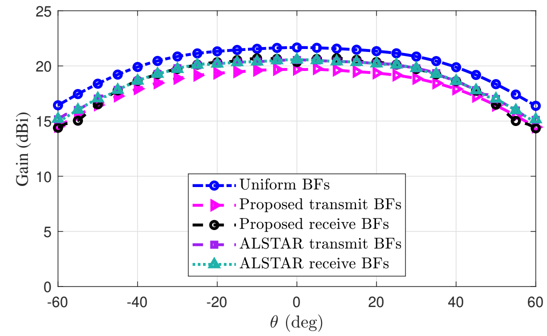

Fig. 4. Beamforming gains for uniform BFs, the proposed transmit BFs, the proposed receive BFs, ALSTAR transmit BFs and ALSTAR receive BFs over $\pm 60^\circ$ beam scan angles (ϕ is 90°).

in Fig. 2. First, compared with uniform BFs without SIC techniques, the proposed method and ALSTAR method both achieve significant EII improvement about 92 dB that is close to the upper bound for total transmitted power $P_\mathrm{t} = 360$ W at broadside under perfect SI channel estimation. However, as about 66 dB EII improvement is achieved by the digital cancellation filters alone, this reveals that the proposed method can provide the same near-upper bound EII performance as the ALSTAR method while avoiding the need for additional SIC techniques that increase system complexity. Second, for different ϵ, the proposed method exhibits robustness against the estimation error. For $\epsilon = 10^{-2}$ and 10^{-3}, the proposed method still provides considerable increase of EII performance than the uniform BFs and even provides better performance than uniform BFs with digital cancellation filters for $\epsilon = 10^{-3}$.

However, the total isolation of EII does not reflect SIC performance before the ADC in each Rx channel. Therefore, Fig. 3 shows the ADC bit loss of the twelve most saturated Rx channels that are adjacent to the Tx subarray for ALSTAR transmit BFs and the proposed transmit BFs. Compared with ALSTAR transmit BFs, the proposed method achieves much lower bit loss in all these Rx channels and at all different P_t levels, which demonstrates the effectiveness of the proposed per-antenna transmit BF design that achieves more isolation on all Rx channels. As the most saturated Rx channel in Rx subarray is one of these twelve adjacent Rx channels, this result also indicates that the proposed method can preserve more dynamic range of all ADCs in Rx subarray. Meanwhile, the result demonstrates that the proposed method has potential to be implemented in digital phased arrays for low power scenarios with lower ADC dynamic range requirement and in higher power scenarios with some additional ADC dynamic range. On the contrary, after ALSTAR transmit BF design, the adjacent Rx channels might be still saturated. Therefore, the proposed method achieves more isolation on each Rx channel before ADC and preserves more dynamic range of ADC to better enable STAR in digital phased arrays.

Then, the aperture performance compromise of the proposed method in beamforming gains are shown in Fig. 4. Compared with the uniform BFs, the Tx and Rx gains of the proposed method are reduced by about 2 dB and 1 dB nearby the array normal, and the gain losses are about 2 dB and 2 dB around the maximum scan angle of $\pm 60^\circ$, respectively. The Rx gain of the proposed method is basically the same as that of ALSTAR method. The Tx gain of the proposed method is slightly lower than that in ALSTAR between $\pm 60^\circ$ scan angles, which is the tradeoff for more isolation before Rx channels and removal of the digital cancellation filters. This tradeoff is generally acceptable while achieving high isolation with lower system complexity. Meanwhile, this result also demonstrates the effectiveness of the tradeoff design in transmit BFs, as γ is set to be 2 dB lower than the Tx gain of uniform transmit BF.

The computational complexity of the proposed method and the comparison with ALSTAR are briefly discussed. For the proposed method, it is about $\mathcal{O}(T_1 M^{3.5} + N^3 + (N^2 M + M^2))$. For ALSTAR, it is about $\mathcal{O}(T_2 T_3 (M^3 + N^3 + (N^2 M + NM^2)))$, where T_2 is the iteration number and T_3 is the number of different initial weights for better local optimum convergence. In practice, due to the well-formulated problem structure and the acceleration from the solver, the proposed method shares similar time complexity with ALSTAR to calculate a set of transmit and receive beamformers.. Moreover, if T_3 is considered by ALSTAR, the proposed method will outperform ALSTAR in complexity.

V. Conclusion

In this letter, we have presented a SIC beamforming design for digital phased arrays that only utilizes DBF technique with lower system complexity and consider practical constraints on ADCs to achieve more isolation before ADC in each Rx channel. The transmit SIC design exploits a per-antenna scheme that preserves more dynamic range for ADC in each Rx channel. The proposed adaptive receive BFs further mitigates the residual SI and noise to achieve high total isolation. The numerical results demonstrate the effectiveness and superiority of the proposed method at different power levels in terms of bit loss in each Rx antenna, achieving near-upper bound SIC performance of EII, and adjustable losses in beamforming gains. In our future work, the proposed method will be applied in different scenarios to further verify its effectiveness.

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