Secure Hybrid Analog and Digital Receive Beamforming Schemes for Efficient Interference Reduction

Yaolu Qin, Shuping Yang, Xiaobo Zhou, Jinhui Lu, Yijin Zhang, and Jiangzhou Wang

Abstract—In spatial spectrum estimation field, medium-scale or large-scale receive antenna array with digital beamforming can be applied at receiver to achieve a high-resolution direction of arrival (DOA) estimation and make a significant interference suppression, but leads to an expensive RF-chain circuit cost. Thus, a robust hybrid analog-and-digital beamforming (ADB) structure is proposed to greatly reduces the number of RF-chains so that the cost of RF-chains is lowered significantly. This ADB structure at receiver is to attenuate the jamming signals from undesired directions and boost the receive quality of the desired signal for the purpose of secure receiving. In this paper, we first propose a hybrid ADB scheme with analytic expression, which consists of null space projection in analog beamforming part and diagonal loading method in digital beamforming part. Then, by exploiting DOA estimation errors, we construct a robust hybrid beamforming algorithm using the conditional expectation of DOA instead of making direct use of the estimated DOA. Simulation result shows that the proposed robust beamforming scheme performs more robust and makes a more efficient reduction in jamming interference to improve the receive signal to interference and noise ratio compared to non-burst ADB scheme proposed by us.

Index Terms—Hybrid beamforming, secure, robust, energy-efficient.

I. INTRODUCTION

With the rapid development of wireless communication and mobile networks, security problem in wireless networks has attracted more and more research activities from both academia and industry, and is now becoming one of the most important problems in wireless networks [1], [2], [3], [4], [5], [6]. In [1], the author’s pioneer research work has laid a theoretical and model foundation for physical layer security. In addition to conventional encryption technology employed at the higher protocol layer for secure data transmission, physical layer security [3] provides another-layer protection for confidential messages. Most recent secure research work focuses on how to design the precoding vector and artificial noise matrix at transmitter in accordance with some given criteria. Direction modulation, as an attractive secure transmit way, preserves the original signal constellation of transmitted signals along the desired direction unchanged, and distorts the signal constellation in the undesired direction. In [7], the authors proposed an orthogonal vector approach, using the method of null space projection (NSP), which improves the information secrecy. The measurement error of direction angle was taken into account in [8], [9], [10]. In [8], a novel and robust direction modulation synthesis method based on conditional minimum mean square error (MMSE) was proposed to alleviate the performance degradation due to the angle measurement error. In [9], the authors extended the idea of [8] to the broadcasting scenario with multiple desired users. Moreover, in [10], a blind robust leakage method of mainlobe-integration was especially designed for multi-user MIMO scenario when direction measurement errors are unknown.

However, at receiver, passive and active source interference is also a serious secure issue. For global navigation satellite systems, jamming and spoofing from hostile emitter or transmitter will seriously degrade its performance. In [11], a two-step beamforming scheme was proposed to both suppress interference and deception by first constructing orthogonal subspaces of spoofing signals and then anti-jamming by cross spectral self-coherence restoral algorithm. Similar to global navigation satellite systems, it is also very easy to disturb synthetic aperture radar and high-frequency ground wave radar. In [12], the authors proposed a novel interference suppression algorithm using robust principal component analysis based signal separation in time-frequency domain to address the narrow-band interference of synthetic aperture radar. In [13], by constructing interference subspace at the reserved range bins and projecting the echoes onto its orthogonal subspace at the interested range bins, the extraction precision of sea state parameters would be improved. In [14] and [15], the MMSE was adopted as an effective way to suppress interference. Some new MMSE interference suppression schemes were proposed in [14], which can be implemented adaptively when interference parameters are unknown or time-varying. In [15], the authors developed a MMSE-based method and recursive least squares (RLS) based scheme with the latter requiring much less computational complexity than the former.

Massive multiple-input multiple-output (MIMO) system, in which employs multiple transmit and receive antennas, has emerged to achieve both extremely high reliability and spectral efficiency. However, in digital beamforming structure, each antenna needs one radio frequency (RF) chain, which will require a large hardware cost when array size tends to large scale. For a massive MIMO system, a hybrid precoding structure is a natural choice to lower the high circuit cost [16]. In [17], a mixed analog-to-digital converter (ADC) receiver architecture was proposed, combining costly high and less...
expensive low resolution ADCs, was worse in performance than the full-resolution ADC structure. Therefore, in [16], a hybrid analog and digital precoding algorithm was firstly proposed to make a balance between hardware cost and system performance. In traditional hybrid beamforming system, the main aim is to maximize the spectral efficiency with known transmit signals. In [13], the authors developed a low-complexity alternating minimization precoder by enforcing an orthogonal constraint on the digital precoder. An energy-efficient hybrid precoding for sub-connected architecture was proposed in [19]. In [20], the authors presented a receive baseband combiner with the target of minimizing mean-squared-error between transmitted and processed received signals. To make a good balance between energy efficiency and spectrum efficiency in [21], the green point for fixed product of the transceivers number and the active antennas number per transceiver was analyzed. Also taking the rate into account, the authors in [22] developed an iterative hybrid beamforming algorithm for the single user in mmWave channel, which can approach the rate limit achieved by unconstrained digital beamforming solutions.

To the best of our knowledge, how to use a hybrid beamforming structure to make an efficient jamming reduction with DOA available at the receiver is an open challenging problem. In this paper, each subarray output of the hybrid structure is viewed as one single virtual large antenna output when we do digital beamforming operation. We will focus on the aspect research and make our effort to address this problem, our main contributions are summarized as follows:

1) With perfect DOA knowledge available, by fully exploiting the sub-array structure, a hybrid non-robust beamforming scheme, a combination of NSP and diagonal loading (DL) algorithm in [23], is proposed, where phase alignment is used in analog part and DL is adopted in digital by adjusting both phase and magnitude. Compared to traditional DL method, the proposed hybrid scheme greatly reduces the RF chains cost and achieves a better jamming reduction.

2) Taking the DOA measurement errors into account, and avoiding the mismatch of array manifold, an improved robust hybrid beamforming scheme is proposed. When the DOA measurement error obeys a uniformly distribution, we use the conditional expectation of the steering vector instead of the steering vector directly generated by the estimated DOA. Simulation results show that the improved robust beamforming scheme is verified to be robust against the DOA measurement errors compared with existing non-robust methods.

The remainder of this paper is organized as follows. Section II describes our system model. In Section III, a non-robust hybrid beamforming scheme of combining NSP and DL is proposed to dramatically reduce the cost of RF chains. Subsequently, in Section IV, an improved robust hybrid beamforming scheme of combating the DOA measurement errors is proposed. Section V presents simulation results to evaluate the performance of our proposed algorithms. Finally, conclusions are drawn in Section VI.

II. System Model

To improve energy efficiency and reduce circuit cost [19], Fig. 1 sketches a sub-array-connected receive structure, one of two typical ABD structures, where each antenna is connected to one phase shifter. Below, by using this structure, the interference signals with DOAs different from the desired direction will be compressed to guarantee secure transmission. In Fig. 1, one desired emitter and Q interference emitters transmit the following narrow band signals: $s_d(t)e^{j2\pi f_d t}$, and $s_1(t)e^{j2\pi f_1 t}, \cdots, s_Q(t)e^{j2\pi f_Q t}$, where $s_d(t)$ is the baseband signal of our desired signal and $s_1(t), \cdots, s_Q(t)$ are the baseband signals of Q jamming signals. They operate on the same frequency band, and so Q interference signals are the co-channel interference (CCI) for the receiver. Considering the CCI, all received signals are working at the same frequency. The linear uniform array (ULA) consists of N omnidirectional antenna elements and is divided into K subsets, with each subset has M antennas, i.e., $N = KM$. Without loss of generality, it is supposed that the desired signal and interference signals are coming from $Q + 1$ different DOAs as follows

Notation: throughout the paper, matrices, vectors, and scalars are denoted by letters of bold upper case, bold lower case, and lower case, respectively. Signs $(\cdot)^T$ and $(\cdot)^H$ denote transpose and conjugate transpose, respectively. Notation $E\{\cdot\}$ stands for the expectation operation. Matrices $I_N$ denotes the $N \times N$ identity matrix and $0_{M \times N}$ denotes $M \times N$ matrix of all zeros.
we have the following baseband signal vector 
\[ \tilde{y}_k(t) = \sum_{m=1}^{M} s_d(t) e^{(2\pi f_c t - 2\pi f_c \tau_d - \alpha_{k,m})} + \sum_{q=1}^{Q} \sum_{m=1}^{M} s_q(t) e^{(2\pi f_c t - 2\pi f_c \tau_q - \alpha_{k,m})} + n_k(t), \]

where \( \tau_d \) and \( \tau_q \) are the propagation delays determined by the direction of the received signals relative to the array, and given by

\[ \tau_d = \tau_0 - \frac{d_m}{c} \sin \theta_d, \]

and

\[ \tau_q = \tau_0 - \frac{d_m}{c} \sin \theta_q, \]

where \( \tau_0 \) is the propagation delay from the emitter to the reference point on the array, \( c \) is the speed of light, and \( d_m \) are the distances of the array elements from the reference point. In this paper, it is assumed that the first element is the reference point, and \( d \) denotes the antenna spacing. \( \tau_0 \) and \( \tau_0 \) is rewritten as

\[ \tau_d = \tau_0 - \frac{(k-1)M + m - 1}{c} \sin \theta_d, \]

and

\[ \tau_q = \tau_0 - \frac{(k-1)M + m - 1}{c} \sin \theta_q. \]

In \( \alpha_{k,m} \) is the corresponding phase for analog beamformer \( \mathbf{W}_{RF} \) correspond to \( m \)th antenna of subarray \( k \). Stacking all \( K \) subarray outputs in \( \tilde{y}(t) \) forms the following matrix-vector notation

\[ \tilde{y}(t) = e^{j2\pi f_c t} \mathbf{W}_{RF}^H \mathbf{A} s(t) + \mathbf{n}(t), \]

where \( s(t) = [s_1(t), s_1(t), \ldots, s_Q(t)]^T \) and \( \mathbf{n}(t) = [n_1(t), n_2(t), \ldots, n_K(t)]^T \) is an additive white Gaussian noise (AWGN), whose entries are independent identically distributed. \( \mathcal{CN}(0, \sigma_n^2) \), and the steering matrix \( \mathbf{A} \) is defined by

\[ \mathbf{A} = [\mathbf{a}(\theta_1), \mathbf{a}(\theta_2), \ldots, \mathbf{a}(\theta_Q)], \]

where \( \mathbf{a}(\theta) \) is the so-called array manifold

\[ \mathbf{a}(\theta) = [1, e^{j2\pi d \sin \theta}, \ldots, e^{j2\pi (N-1)d \sin \theta}]^T, \]

and the \( \mathbf{W}_{RF} \) is \( N \times K \) phase shift matrix,

\[ \mathbf{W}_{RF} = \begin{bmatrix}
\mathbf{f}_1 & 0 & \cdots & 0 \\
0 & \mathbf{f}_2 & \cdots & 0 \\
\vdots & \vdots & \ddots & \vdots \\
0 & 0 & \cdots & \mathbf{f}_K
\end{bmatrix}, \]

where \( \mathbf{f}_k = \frac{1}{\sqrt{M}} [e^{j\alpha_{1,k}}, e^{j\alpha_{2,k}}, \ldots, e^{j\alpha_{M,k}}]^T \) is the analog beamforming vector of the \( k \)th subarray. The radio frequency (RF) signal \( \tilde{y}(l) \) in \( \mathbf{W}_{RF} \) passes through \( K \) parallel RF chains, which contain the corresponding down converters and ADCs, we have the following baseband signal vector

\[ \tilde{y}(l) = \mathbf{W}_{RF}^H \mathbf{A} s(l) + \mathbf{n}(l), \]

Via digital beamforming operation, the above signal vector becomes

\[ r(l) = \mathbf{w}_{BB}^H \mathbf{W}_{RF}^H \mathbf{A} s(l) + \mathbf{w}_{BB}^H \mathbf{n}(l), \]

where \( \mathbf{w}_{BB} = [w_1, w_2, \ldots, w_K]^T \) stands for the digital beamformer.

### III. Proposed Non-robust Hybrid Beamforming Scheme

In this section, given a perfect DOA estimate, a non-robust hybrid ADB beamforming scheme is proposed to achieve a significant interference reduction compared to DL in fully-digital structure.

#### A. Design of Total Beamforming Vector

The total steering matrix \( \mathbf{A} \) is expressed as \( [\mathbf{a}(\theta_1), \mathbf{A}_I] \), where \( \mathbf{A}_I \) composes all steering vectors of interference signals. Firstly, let us define the total beamforming vector \( \mathbf{W}_{RF} \mathbf{w}_{BB} = \mathbf{v} \), which is viewed as one single optimization variable. From the rule of NSP, the optimization problem is casted as

\[ \begin{aligned}
\text{maximize} & \quad ||\mathbf{v}^H \mathbf{a}(\theta_d)||^2 \\
\text{subject to} & \quad \mathbf{A}_I^H \mathbf{v} = 0,
\end{aligned} \]

In accordance with the above equality constraint, the total beamforming vector \( \mathbf{v} \) is orthogonal to the null space of \( \mathbf{A}_I^H \). To construct \( \mathbf{v} \), the singular value decomposition (SVD) operation is performed on conjugate transpose interference matrix \( \mathbf{A}_I^H \) as follows: \( \mathbf{A}_I^H = \mathbf{U} \mathbf{\Sigma} \mathbf{V}^H \), where \( \mathbf{U} \) and \( \mathbf{V} \) are the unitary matrices with \( \mathbf{U} \in \mathbb{C}^{d \times Q} \), and \( \mathbf{V} \in \mathbb{C}^{N \times N} \). \( \mathbf{\Sigma} \in \mathbb{C}^{Q \times N} \) is a rectangle matrix with singular values on its main diagonal and all off-diagonal elements are zero, i.e. \( \sum \mathbf{\Sigma} = [\text{diag} \{\sigma_1^2, \sigma_2^2, \ldots, \sigma_Q^2\}, \mathbf{0}_{Q \times (N-Q)}] \).

According to the equality constraint in \( \mathbf{v} \) is given by a linear combination of the \( N - Q \) most right columns of \( \mathbf{V} \), i.e.

\[ \mathbf{v} = \mathbf{F} \mathbf{\tilde{v}}, \]

where \( \mathbf{F} \) is the \( N - Q \) most right columns of \( \mathbf{V} \), and \( \mathbf{\tilde{v}} \) is a column vector with each entry controlling the linear combination of right singular vectors and it’s normalized power \( E(\mathbf{\tilde{v}}^H \mathbf{\tilde{v}}) = 1 \).

Therefore, the optimization problem in \( \mathbf{v} \) can be rewritten as

\[ \begin{aligned}
\text{maximize} & \quad ||\mathbf{\tilde{v}}^H \mathbf{F}^H \mathbf{a}(\theta_d)||^2, \\
\text{subject to} & \quad ||\mathbf{F}^H \mathbf{a}(\theta_d)|| = 1,
\end{aligned} \]

which directly yields

\[ \begin{aligned}
\mathbf{\tilde{v}}_{opt} &= \mathbf{F}^H \mathbf{a}(\theta_d) / ||\mathbf{F}^H \mathbf{a}(\theta_d)||, \\
\mathbf{v}_{opt} &= \mathbf{F} \mathbf{\tilde{v}}_{opt}.
\end{aligned} \]
B. DL-based Digital Beamformer

In what follows, given the initial value of analog beamformer \( \mathbf{W}_{RF_0} \), we show how to optimize the digital beamforming vector \( \mathbf{w}_{BB} \). A wise choice is to make the array point towards the DOA of the desired signal, i.e.,

\[
\alpha_{k,m,0} = \frac{2\pi}{\lambda} ((k-1)M + m - 1) \sin \theta_d. \tag{17}
\]

where \( \alpha_{k,m,0} \) is the initial corresponding phase of \( \mathbf{W}_{RF_0} \). The steering vector of subarray \( \mathbf{a}_{sub} \) is generated with known \( \mathbf{W}_{RF_0} \), let us define the subarray steering vector of the desired signal as follows

\[
\mathbf{a}_{sub}(\theta_d) = \mathbf{W}^H_{RF_0} \mathbf{a}(\theta_d), \tag{18}
\]

The perfect information of noise and signals cannot be available in practice, therefore the sampling covariance matrix is adopted instead. Since the Capon beamforming method is sensitive to modeling errors, it is not robust for modeling mismatch, which will lead output of Capon beamformer deteriorate seriously. Therefore, here, instead of Capon method, the DL method is employed to design digital beamforming vector, which is robust to modeling mismatch. Via the regularization operation on Capon method, the optimization problem of using DL method to optimize the beamforming vector \( \mathbf{w}_{BB} \) can be casted as

\[
\begin{align*}
\text{minimize} & \quad \mathbf{w}^H_{BB} \left( \hat{\mathbf{R}} + \gamma \mathbf{I} \right) \mathbf{w}_{BB} \\
\text{subject to} & \quad \mathbf{w}^H_{BB} \mathbf{a}_{sub}(\theta_d) = 1,
\end{align*}
\]

where \( \gamma \) denotes the DL factor, and \( \hat{\mathbf{R}} \) is the sampling covariance matrix corresponding to \( K \) subarrays,

\[
\hat{\mathbf{R}} = \frac{1}{L} \sum_{l=1}^{L} \mathbf{y}(l) \mathbf{y}^H(l). \tag{20}
\]

where \( L \) is the number of snapshots. Applying the Lagrange multiplier method to the optimization problem \( \text{[19]} \), the associated Lagrangian function has the following form

\[
f(\mathbf{w}_{BB}, \lambda) = \mathbf{w}^H_{BB} \left( \hat{\mathbf{R}} + \gamma \mathbf{I} \right) \mathbf{w}_{BB} + \lambda \left( \mathbf{w}^H_{BB} \mathbf{a}_{sub}(\theta_d) - 1 \right), \tag{21}
\]

where the scalar \( \lambda \) is the lagrange multiplier. By taking the derivation of \( f(\mathbf{w}_{BB}, \lambda) \) with respect to \( \mathbf{w}_{BB} \) and making it 0, the digital beamformer is

\[
\mathbf{w}_{BB} = -\lambda \left( \hat{\mathbf{R}} + \gamma \mathbf{I} \right)^{-1} \mathbf{a}_{sub}(\theta_d). \tag{22}
\]

Substitute \( \text{[22]} \) into the equation constraint of \( \text{[19]} \), then \( \mathbf{w}_{BB} \) is expressed as

\[
\mathbf{w}_{BB} = \frac{\left( \hat{\mathbf{R}} + \gamma \mathbf{I} \right)^{-1} \mathbf{a}_{sub}(\theta_d)}{\mathbf{a}^H_{sub}(\theta_d) \left( \hat{\mathbf{R}} + \gamma \mathbf{I} \right)^{-1} \mathbf{a}_{sub}(\theta_d)}. \tag{23}
\]

Observing the above expression, the main advantage of the DL method ensures that the diagonal loading matrix \( \hat{\mathbf{R}} + \gamma \mathbf{I} \) is invertible by adding white noise to diagonal elements of sampling covariance matrix \( \hat{\mathbf{R}} \).

C. Analytic Analog Beamformer

Now, we turn to the construction of the analog beamformer \( \mathbf{W}_{RF} \). To reduce interference and maximize received power of desired signal, we model the problem of optimizing \( \mathbf{W}_{RF} \) as follows

\[
\begin{align*}
\text{minimize} & \quad \| \mathbf{v}_{opt} - \mathbf{W}_{RF} \mathbf{w}_{BB} \| \\
\text{subject to} & \quad \| \mathbf{v}_{opt} - \mathbf{W}_{RF} \mathbf{w}_{BB} \| = \frac{1}{\sqrt{M}} e^{j\theta_m} w_k,
\end{align*}
\]

which directly yields the following closed-form solution

\[
\alpha_{k,m} = \angle \left( \frac{\mathbf{v}_{opt}(k-1)\cdots M + m}{w_k} \right), \tag{26}
\]

where \( k \in \{1, 2, \cdots, K\}, m \in \{1, 2, \cdots, M\}, \) and \( \mathbf{v}_{opt} \) and \( w_k \) represents the \( k \)th element of \( \mathbf{v}_{opt} \) and \( \mathbf{w}_{BB} \), respectively. Based on the above construction, we summarize our non-robust hybrid algorithm as the Algorithm I.

Algorithm 1 Non-robust Hybrid Beamforming Algorithm Based on DL Method via NSP

Input: \( \mathbf{W}_{RF_0} \)

1: Calculate \( \mathbf{v}_{opt} \) by \( \text{[16]} \);

2: According to initial analog beamforming matrix, compute \( \mathbf{a}_{sub}(\theta_d) \) by \( \text{[18]} \);

3: DL-based digital beamforming vector \( \mathbf{w}_{BB} \) is constructed in accordance with \( \text{[23]} \);

4: Reconstruct \( \alpha_{k,m} \) by \( \text{[26]} \);

Output: \( \mathbf{W}_{RF}, \mathbf{w}_{BB} \)

IV. PROPOSED ROBUST HYBRID BEAMFORMING SCHEME

In practical applications, we can only obtain the estimated value of DOA. If we have the prior knowledge of DOA measurement errors. By fully exploiting the statistical knowledge of DOA measurement errors, the robust hybrid beamforming scheme is proposed to achieve an obvious interference reduction in the following compared to the hybrid one in the previous section.

A. Design of Robust Analog Beamformer \( \mathbf{W}_{RF} \)

In the presence of DOA measurement errors, the ideal DOA can be modelled as

\[
\theta = \hat{\theta} + \Delta \theta, \tag{27}
\]

where \( \hat{\theta} \) denotes the estimated DOA, and \( \Delta \theta \) is the angle error. Using \( \text{[27]} \), the optimization problem in \( \text{[12]} \) is rewritten as

\[
\begin{align*}
\text{maximize} & \quad \| \mathbf{v}^H \mathbf{a}(\hat{\theta}_d + \Delta \theta_d) \|^2 \\
\text{subject to} & \quad \mathbf{A}_f(\hat{\theta}_q + \Delta \theta_q) \mathbf{v} = \mathbf{0}, \quad q \in \{1, 2, \cdots, Q\},
\end{align*}
\]
In this paper, it is assumed that $\Delta \theta$ is uniformly distributed over the interval $[-\varepsilon, \varepsilon]$. Let $p(\Delta \theta)$ denotes the probability distribution of $\Delta \theta$.

$$p(\Delta \theta) = \begin{cases} \frac{1}{2\varepsilon} & -\varepsilon \leq \Delta \theta \leq \varepsilon \\ 0, & \text{otherwise}, \end{cases}$$

(29)

where $\varepsilon$ is maximum error between estimated DOA $\hat{\theta}_d$ and exact DOA $\theta_d$. Due to the error of DOA estimation error $\Delta \theta$, the exact DOA $\theta = \hat{\theta} + \Delta \theta$ can also be viewed as uniformly distributed. Correspondingly, the expectation of exact DOA

$$E[r] = \int_{-\varepsilon}^{\varepsilon} e^{j \frac{2\pi}{\lambda} (i-1) d \sin (\hat{\theta}_d + \Delta \theta_d)} \times p(\Delta \theta)d(\Delta \theta_d)$$

(31)

In order to simplify (31), let us define

$$a_i = j \frac{2\pi}{\lambda} (i-1) d \sin \hat{\theta}_d$$

(32)

$$b_i = j \frac{2\pi}{\lambda} (i-1) d \cos \hat{\theta}_d$$

$$c \triangleq \frac{\varepsilon}{\pi}.$$

Therefore

$$r_i = \frac{1}{2\pi} \int_{-\pi}^{\pi} e^{a_i x + b_i x} dx.$$

(33)

Similar to the derivation of $E[\mathbf{a}(\theta_d)]$, we have

$$E[\mathbf{a}_R^H(\theta_q)] = E[\mathbf{a}_R^H(\hat{\theta}_q + \Delta \theta_d)] \triangleq \mathbf{R},$$

(34)

where $q \in \{1, 2, \cdots, Q\}$. The entry at the $i$th row and the $q$th column

$$R_{iq} = \frac{1}{2\pi} \int_{-\pi}^{\pi} e^{a_{iq} x + b_{iq} x} dx,$$

(35)

where

$$a_{iq} \triangleq j \frac{2\pi}{\lambda} (i-1) d \sin \hat{\theta}_q$$

(36)

$$b_{iq} \triangleq j \frac{2\pi}{\lambda} (i-1) d \cos \hat{\theta}_q$$

$$c \triangleq \frac{\varepsilon}{\pi}.$$

Substituting $\mathbf{r}$ and $\mathbf{R}$ into (28) forms the following robust optimization problem

$$\begin{align*}
\text{maximize} & \quad ||\mathbf{v}^H\mathbf{r}||^2 \\
\text{subject to} & \quad \mathbf{R}^H\mathbf{v} = 0.
\end{align*}$$

(37)

Similar to Section III-A, the total robust beamformer $\mathbf{v}$ is derived by making use of the above optimization as follows

$$\mathbf{v}_{\text{opt}} = \frac{\hat{\mathbf{F}}^H\mathbf{r}}{||\hat{\mathbf{F}}^H\mathbf{r}||}.$$

(38)

where $\hat{\mathbf{F}}$ is the $N - Q$ most right columns of singular vector of $\mathbf{R}$.

Similar to the derivation of $\mathbf{r}$ and $\mathbf{R}$, in order to calculate the initialization value $\mathbf{W}_{RF,0}$, the expectation of $\sin \theta_d$ is adopted to replace real $\sin \hat{\theta}_d$, i.e.,

$$E[\sin \theta_d] = E[\sin (\hat{\theta}_d + \Delta \theta_d)]$$

$$= \int_{-\varepsilon}^{\varepsilon} \sin(\hat{\theta}_d + \Delta \theta_d) \times p(\Delta \theta)d(\Delta \theta_d)$$

(39)

$$= \frac{1}{\varepsilon} \sin \hat{\theta}_d \sin \varepsilon.$$

Thus, the corresponding phase of $\mathbf{W}_{RF,0}$ is given by

$$\alpha_{k,m,0} = \frac{2\pi}{\varepsilon \lambda} ((k - 1)M + m - 1) \sin \hat{\theta}_d \sin \varepsilon.$$

(40)

B. Robust Digital Beamformer $\mathbf{w}_{BB}$ Design

In order to improve the robustness of beamforming in digital part, we modify the constraints of DL method so that signal can pass without distortion with the angle around estimated DOA, which is modelled as the following optimization problem

$$\begin{align*}
\text{minimize} & \quad \mathbf{w}_{BB}^H (\hat{\mathbf{R}} + \gamma \mathbf{I}) \mathbf{w}_{BB} \\
\text{subject to} & \quad \mathbf{w}_{BB}^H \mathbf{a}_{\text{sub}}(\hat{\theta}_d + \Delta \theta_d) = 1,
\end{align*}$$

(41)

According to (23) with perfect DOA estimation, when there exists DOA estimation error $\Delta \theta_d$, the DL beamformer is directly defined as

$$\mathbf{w}_{BB} = \frac{(\hat{\mathbf{R}} + \gamma \mathbf{I})^{-1} \mathbf{W}_{RF,0}^H \mathbf{r}}{\mathbf{R}^H \mathbf{W}_{RF,0} (\hat{\mathbf{R}} + \gamma \mathbf{I})^{-1} \mathbf{W}_{RF,0}^H \mathbf{r}}.$$

(42)

V. SIMULATION RESULTS

In this section, we present simulation results to assess the performance of non-robust hybrid beamforming and improved robust hybrid beamforming proposed by us. Simulation parameters are set as follows: $N = 32$, $K = 4$, and antenna spacing $d = 0.5\lambda$. It is assumed that the estimated desired DOA $\theta_d = 60^\circ$, and the estimated DOAs of interference sources are $30^\circ$ and $-15^\circ$. 
Firstly, with perfect DOA knowledge available, Fig. 2 shows the curves of beamforming gain versus direction of our proposed non-robust hybrid algorithm. Apparently, the proposed method can attenuate the beam gain by at least $-30\,\text{dB}$ in the direction of interference signals, where signal to noise ratios (SNR) of desired and interference sources are 0dB and 15dB, respectively. In contrast with traditional method, it performs better along the direction of interference and has the same gain in the desired direction.

In Fig. 3 we presents the curves of signal to interference plus noise ratio (SINR) versus SNR of the desired signal by ranging SNR from -15dB to 15dB. From this figure, it is seen that the proposed hybrid beamformer almost has the same performance as DL method in the low SNR region. With increasing signal power, our algorithm has a stronger ability to reduce interference. Specifically, when the SNR of the desired signal is less than -5dB, the SINR curve of our proposed algorithm and DL method are identical, however, when SNR is higher than -5dB, the SINR curve of our proposed algorithm is above that of DL. Therefore, it is substantially reasonable to adopt the non-robust hybrid algorithm with exact DOA estimation as a reference to evaluate the performance of robust algorithm.

According to the estimator presented by [25], the root mean square (RMS) is less than $1^\circ$ under the condition that the SNR of estimated source is 0dB. In order to examine the robust performance of our proposed method in extreme conditions, it is supposed that $\Delta \theta$ is uniformly distributed and there exists maximum estimation error $3^\circ$, i.e., $\varepsilon = 3^\circ$.

In Fig. 4 demonstrates the curves of SINR versus SNR for our proposed methods, where the maximum estimate error is $3^\circ$. Observing this figure, we find, with the increase in the value of SNR, the SINR performance of non-robust hybrid beamforming with DOA estimation errors is always almost 3dB worse than that with perfect DOA estimation. It is noted that when the desired signal power is low, the improved robust algorithm is closer to the curve with perfect DOA, that is, our improved robust hybrid beamforming can reduce interference and retain robust when there exists direction-finding errors in the low SNR region, which is the typical situations of interference reduction.

Fig. 5 plots the curves of root-mean-square-error (RMSE) versus maximum measurement angle errors for the proposed non-robust hybrid beamforming with perfect DOA used as a reference. By our calculation, when there exist DOA estimation errors, the RMSE differences between the reference and three methods grow gradually with angle error. It is noted that, with the increase of $\varepsilon$, the RMSEs of non-robust hybrid beamforming, DL method and robust hybrid beamforming all become worse. However, the RMSE of robust hybrid beamforming and DL algorithm are always lower than that...
SNR of desired signal is 0 dB

Fig. 5. Deviations between non-robust hybrid beamforming without DOA estimation error and three methods with DOA estimation errors versus maximum angle error.

of non-robust hybrid beamforming, which means it has better performance with different $\varepsilon$.

Fig. 6. SINR versus snapshot number for 3 different SNRs of desired signal

At last, Fig. 6 illustrates the effect of snapshot number on the robust hybrid beamforming scheme proposed by us. From Fig. 6 it is shown that the performance of the proposed robust hybrid beamforming is gradually improved with increasing the snapshot number. When the receive SNR of the desired signal is in the low SNR region, the number of snapshots has more important impact on the performance improvement. When the number of snapshots reaches the limit value 32, it has little impact on performance improvement.

VI. CONCLUSION

In this paper, two hybrid beamforming schemes, non-robust and robust, have been proposed, which are based on DL method, where the NSP rule is adopted to design the total beamforming vector. In the case of perfect DOA available, the proposed non-robust hybrid scheme achieves a substantial beam gain over DL algorithm in the desired and interference directions. In the presence of DOA measurement errors, the proposed robust hybrid beamforming, with uniformly distributed angle errors, shows a good robustness compared to non-robust method DL and non-robust hybrid method. As the maximum angle error becomes larger, its SINR performance over non-robust schemes such as DL becomes more significant.

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