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Research Article

Robust Mainlobe Interference Suppression Based on Joint Oblique Projection for Dual Polarization Conformal Array

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In this paper, a robust mainlobe interference suppression algorithm is proposed in a space-polarization joint domain based on a dual-polarized conformal array. An oblique projection filter can eliminate the interference while retaining the desired signal information. However, little deviation in signal parameters will lead to significant performance degradation. To solve such problems, the joint steering vector of the desired signal is corrected by solving the quadratic convex optimization problem. Parameters of interferences are first estimated by using multiple signal classification (MUSIC) joint spectrum, and then, the interference covariance matrix can be reconstructed individually to further estimate the interference steering vector. After that, the joint oblique projection operator can be established by the estimated steering vectors. Consequently, an antijamming beamformer can be obtained by the joint oblique projection filter to achieve robustness on the interference parameter estimation deviation and desired signal mismatch. Simulation experiments verify the effectiveness of the proposed method.

1. Introduction

Facing the current increasingly complex and changeable electromagnetic environment, main lobe interference is in great demand of array signal processing. Plenty of academics have devoted themselves to researching how to enhance the antijamming capability of the antenna array to improve the performance of receiving signal.

The main lobe interference suppression is a great challenge in traditional adaptive digital beamforming (ADBF) [1]. Blocking Matrix Processing (BMP) [2–4] is proposed to solve the main lobe distortion problem. However, the algorithm needs accurate prior interference information, and it will lead to the problem of missing degrees of freedom. The method of eigen-projection matrix process (EMP) [5, 6] uses the orthogonal characteristics of the signal subspace to construct a pre-processing matrix. Nevertheless, it has the problem of main beam pointing offset, although these methods can suppress the main lobe interference in spatial domain. When the main lobe interfering signal is close to the desired signal, the antijamming performance of these methods will drop sharply.

Several research results on polarized antenna array technology have been published in past years. Literature [7] proposed a compressive sampling framework for two-dimensional direction-of-arrival (DOA) and polarization estimation in millimeter wave polarized massive multiple-input multiple-output (MIMO) systems. In [8], the MIMO radar equipped with polarized array antennas is used to propose a novel framework for unmanned aerial vehicle three-dimensional positioning. With the development of polarized antenna technology, polarization information is introduced for interference suppression. Even if the interference and the desired signal come from close spatial angles, as long as their polarization information is different, the interference can be suppressed with the guaranteed desired signal.

Literature [9] proposed an oblique projection polarization filter (OPPF) based on signal polarization subspace. The filter suppresses interference without destroying the desired signal phase and amplitude information. But only one main lobe interference can be suppressed. Literature [10] used the particle swarm optimization (PSO) algorithm to estimate polarization parameters for polarization filtering to suppress multiple interferences. However, this method has unsatisfactory estimation errors and expensive computational costs. Based on the principle of oblique projection filtering, the method in [11] solves the polarization optimization problem of transmitter/receiver to achieve interference suppression. The previous methods are all based on the accurate parameter estimation of each signal. In practice, it is difficult to estimate polarization parameters without any error. The performance of these methods will decrease significantly when error exists. Literature [12] proposed covariance matrix reconstruction beamforming (CMRB) that extends traditional adaptive beamforming methods to the space-polarization domain based on dual-polarized conformal arrays. By reconstructing the interference covariance matrix in the space-polarization domain, the interference of arbitrary polarization and space status can be suppressed, and it does not rely on precise interference parameters.

Robust adaptive beamforming (RAB) has vital importance for ensuring signal receiving quality. It was analysed in depth and improved by interference-plus-noise covariance matrix (IPNC) reconstruction in many papers [13–20]. In [13], the IPNC and desired signal covariance matrix are reconstructed by estimating all interferences and desired signal power using the principle of maximum entropy power spectrum (MEPS). Literature [14] reconstructs IPNC by using low-complexity spatial sampling process and virtual received array vector. The method in [15] constructs the IPNC based on the idea of using discrete Fourier transform (DFT) to estimate the correlated sequence power spectrum. For literature [16–20], the basic idea of IPNC matrix reconstruction is to estimate the steering vector by maximizing the output power of the beamformer. Literature [16] reconstructs IPNC matrix by estimating the power and the steering vector of the interferences based on the power method. In addition, a desired signal-plus-noise covariance matrix is reconstructed by spatial match processing. In literature [17], the reconstructed IPNC is obtained by Capon integration and then further solved by constructing a convex optimization problem. Literature [18] uses the orthogonality of subspaces to construct the constraints of convex optimization. In literature [19], a subspace is built in a small neighborhood to obtain the steering vector and then search along it to find the Capon power peak. Literature [20] used an interference-plus-noise subspace projection matrix to deal with arbitrary steering vector mismatch. These methods process signals that are all only in the spatial domain and using uniform linear antenna array (ULA) and will lose efficacy when the main lobe interference exists. At present, there is not enough related research on the space-polarization domain, so it is necessary to take further research.

In this paper, the traditional beamforming algorithm is extended to the space-polarization joint domain based on a dual-polarized conformal array. First, aiming at the possible mismatch of the desired signal, a quadratic convex optimization problem is constructed by using the orthogonality between the signal subspace and the noise subspace. Then, the interference covariance matrix is reconstructed individually by integrating in the neighbourhood around the signal. According to subspace properties, the joint steering vector of interference is estimated by eigendecomposition of the reconstructed interference covariance matrix. Finally, the joint oblique projection operator can be established by the estimated signal and interference steering vectors, and consequently, the interference suppression beamformer can be obtained. Simulation experiments show that the proposed method is robust to the estimation error of interference parameters and mismatch of steering vector of desired signal and can effectively eliminate the main lobe interference.

2. Problem Formulation

The dual-polarized antenna can receive and excite the two orthogonal polarization components through two independent feed points. So, dual-polarized antenna can handle signal of arbitrary polarization. A signal with arbitrary polarization state $s(t)$ can be decomposed into two orthogonal polarization components $s_\theta(t)$ and $s_\phi(t)$. We assume that their corresponding steering vectors are $\mathbf{a}_{sp}(\theta,\phi,\gamma,\eta)$, $\mathbf{a}_\theta(\theta,\phi)$ and $\mathbf{a}_\phi(\theta,\phi)$, and then, we have [21]:

$$\mathbf{a}_{sp}(\theta,\phi,\gamma,\eta) = \left[\mathbf{a}_\theta(\theta,\phi), \mathbf{a}_\phi(\theta,\phi)\right]\begin{bmatrix}\sin\gamma e^{j\eta}\\ \cos\gamma\end{bmatrix}, \tag{1}$$

where θ and ϕ represent the elevation angle and azimuth angle and polarization parameters γ and η represent the amplitude ratio and the phase difference.

We consider a desired signal and $M-1$ interference signals with any polarization state impinging on a dual-polarized antenna array. The array that received the signal vector can be expressed as

$$\begin{aligned}\mathbf{x}(t) &= \mathbf{a}_{sp}(\theta_0,\phi_0,\gamma_0,\eta_0)s_0(t)\\ &\quad + \sum_{i=1}^{M-1}\mathbf{a}_{sp}(\theta_i,\phi_i,\gamma_i,\eta_i)s_i(t) + \mathbf{n}(t),\end{aligned} \tag{2}$$

where $s_0(t)$ is the desired signal, $s_i(t)$ is the ith interference, and $\mathbf{n}(t)$ is the independent and identically distributed white Gaussian noise.

In a dual-polarized conformal array with N elements, the polarization signal responds to both orthogonal polarization components of each element. Therefore, all $2N$ channels of received signals need to be jointly processed to obtain optimal weights. The schematic diagram of the beamformer is shown in Figure 1.

Thus, the output of the beamformer can be expressed as

$$y(t) = \mathbf{w}^H\mathbf{x}(t), \tag{3}$$

where $\mathbf{w} = [w_{1\theta}, w_{1\phi}, \cdots, w_{N\theta}, w_{N\phi}]^T$ is the weight vector with the size of $2N\times 1$ for all channels of the conformal array. In addition, the symbol $(\cdot)^H$ stands for Hermitian transpose, and $(\cdot)^T$ is the transpose.

We define the array pattern of the space-polarization domain beamformer as

$$F(\theta,\phi,\gamma,\eta) = \mathbf{w}^H\mathbf{a}_{sp}(\theta,\phi,\gamma,\eta). \tag{4}$$

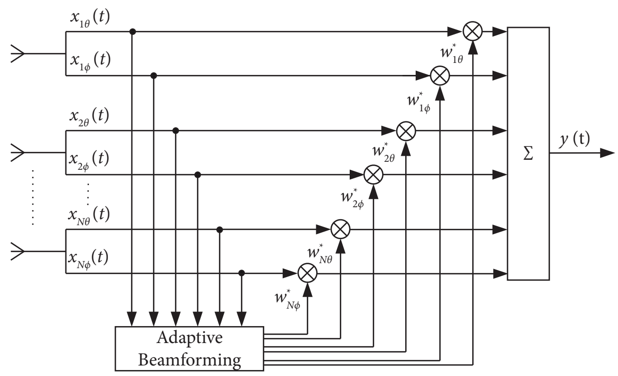

FIGURE 1: Schematic diagram of a dual-polarized array beamformer where $(\cdot)^*$ is conjugate complex.

The signal-to-interference-to-noise ratio (SINR) of the array output can be calculated from

$$\text{SINR} = \frac{\sigma_s^2 \left| \mathbf{w}^H \mathbf{a}_{sp} (\theta_0, \phi_0, \gamma_0, \eta_0) \right|^2}{\mathbf{w}^H \mathbf{R}_{i+n} \mathbf{w}}, \tag{5}$$

where σ_s^2 is the desired signal power, $\mathbf{R}_{i+n}$ is the interference plus noise covariance matrix.

According to the maximum signal-to-interference-to-noise ratio criterion, the traditional minimum variance distortion-less response (MVDR) method is extended to the space-polarization joint domain, which is named SPMVDR in this paper.

$$\begin{aligned} &\min_{\mathbf{w}} \, , \mathbf{w}^{\mathrm{H}} \mathbf{R}_{i+n} \mathbf{w}, \\ &\text{s.t.}, \mathbf{w}^{\mathrm{H}} \mathbf{a}_{sp} (\theta_0, \phi_0, \gamma_0, \eta_0) = 1. \end{aligned} \tag{6}$$

Then, the optimal weight vector $\mathbf{w}_{\text{spmvdr}}$ of SPMVDR can be obtained by solving Equation (6).

$$\mathbf{w}_{\text{spmvdr}} = \frac{\mathbf{R}_{i+n}^{-1} \mathbf{a}_{sp} (\theta_0, \phi_0, \gamma_0, \eta_0)}{\mathbf{a}_{sp}^{\mathrm{H}} (\theta_0, \phi_0, \gamma_0, \eta_0) \mathbf{R}_{i+n}^{-1} \mathbf{a}_{sp} (\theta_0, \phi_0, \gamma_0, \eta_0)}. \tag{7}$$

In practical applications, it is difficult to obtain $\mathbf{R}_{i+n}$.

So, $\mathbf{R}_{i+n}$ is usually replaced by the sampling covariance matrix of the mixed signal $\widehat{\mathbf{R}}$:

$$\widehat{\mathbf{R}} = \frac{1}{L} \sum_{l=1}^{L} \mathbf{x}(l) \mathbf{x}^H (l), \tag{8}$$

where the L is the number of snapshots.

When the signal-to-noise ratio (SNR) of desired signal increases, the performance of the SPMVDR algorithm will decrease. Thus, this paper will explore a new method to solve the problem.

3. Proposed Method

When the interference is close to the desired signal in the spatial domain, it is difficult to suppress the interference while ensuring the reception of the desired signal. Therefore, this paper extends the problem to the space-polarization joint domain by using dual-polarized conformal array. The oblique projection method can completely remove the information of the orthogonal space while retaining the information of the projection space [7]. So, when desired signal and interferences have different space-polarization states, the space-polarization steering vectors of the desired signal and interferences can be used to construct the projection space and the orthogonal space to realize the interference suppression. However, the obtained information of the signals may have errors. Therefore, the steering vectors of desired signal and interferences are first estimated to construct the joint oblique projection operator to enhance the robustness of the proposed method.

3.1. Joint Oblique Projecting Filter. The joint oblique projection filtering operator in space-polarization domain can be defined as

$$\mathbf{Q}_{SV} = \mathbf{S} \left(\mathbf{S}^H \mathbf{P}_V^{\perp} \mathbf{S} \right)^{-1} \mathbf{S}^H \mathbf{P}_V^{\perp}, \tag{9}$$

$$\mathbf{P}_V^{\perp} = \mathbf{I} - \mathbf{V} \left(\mathbf{V}^H \mathbf{V} \right)^{-1} \mathbf{V}^H, \tag{10}$$

where $\mathbf{S}$ and $\mathbf{V}$ are the signal subspace and the interference subspace, $\mathbf{P}_V^{\perp}$ is the orthogonal projection operator of the interference subspace $\mathbf{V}$, and $\mathbf{Q}_{SV}$ is the oblique projection filter operator.

The $\mathbf{S}$ and $\mathbf{V}$ are usually constructed by using the space-polarization steering vectors of signal and interferences, respectively. But the steering vectors of interferences are unknown in most practical applications. They can be calculated by the estimated DOAs and polarization parameters. However, the estimation error of the parameters will lead to significant performance degradation of the joint oblique projecting filter. In addition, the steering vector of desired signal may mismatch which will degrade performance. To solve the problem, the joint steering vectors need to be reestimated to construct $\mathbf{S}$ and $\mathbf{V}$.

3.2. Desired Signal Steering Vector Estimation. In this subsection, the steering vector of desired signal is estimated via solving quadratic convex optimization problem.

The sampling matrix $\widehat{\mathbf{R}}$ can be decomposed as follows:

$$\widehat{\mathbf{R}} = \sum_{i=1}^{M} \lambda_i \mathbf{v}_i \mathbf{v}_i^H + \sum_{i=M+1}^{2N} \lambda_i \mathbf{v}_i \mathbf{v}_i^H, \tag{11}$$

where the eigenvectors $\{\mathbf{v}_1, \cdots, \mathbf{v}_M\}$ correspond to M large eigenvalues form signal subspace $\mathbf{E}_S$ and the remaining eigenvectors $\{\mathbf{v}_{M+1}, \cdots, \mathbf{v}_{2N}\}$ correspond to $2N - M$ small eigenvalues form noise subspace $\mathbf{E}_N$. The noise subspace is the complementary space of the signal subspace, and it is orthogonal to signal subspace.

Set $\bar{\mathbf{a}}_0 = \mathbf{a}_{sp} (\theta_0, \phi_0, \gamma_0, \eta_0)$ represents the presumed steering vector. When the mismatch of steering vector of desired signal exists, the actual steering vector of desired signal can be written as $\mathbf{a}_0 = \bar{\mathbf{a}}_0 + \mathbf{e}$, where $\mathbf{e}$ is the mismatch vector.

Since $\mathbf{e}$ can be decomposed into vectors $\mathbf{e}_{\perp}$ and $\mathbf{e}_{\parallel}$, where $\mathbf{e}_{\perp}$ is perpendicular to $\bar{\mathbf{a}}_0$ and $\mathbf{e}_{\parallel}$ is parallel to $\bar{\mathbf{a}}_0$. It is known that $\mathbf{e}_{\parallel}$ does not affect the performance of beamformer because it is a scaled copy of $\bar{\mathbf{a}}_0$ [17].

Then, the orthogonality of subspaces can be used to obtain the constraint

$$\left(\bar{\mathbf{a}}_0 + \mathbf{e}_{\perp}\right)^{\mathrm{H}} \mathbf{E}_N \mathbf{E}_N^{\mathrm{H}} \left(\bar{\mathbf{a}}_0 + \mathbf{e}_{\perp}\right) = 0. \tag{12}$$

Since constraint (12) is nonconvex and is difficult to be solved, it is adjusted to (13) to ensure that the adjusted orthogonality of $(\bar{\mathbf{a}}_0 + \mathbf{e}_{\perp})$ and noise space $\mathbf{E}_N$ is better than $\bar{\mathbf{a}}_0$.

$$\left(\bar{\mathbf{a}}_0 + \mathbf{e}_{\perp}\right)^{H} \mathbf{E}_N \mathbf{E}_N^{H} \left(\bar{\mathbf{a}}_0 + \mathbf{e}_{\perp}\right) \le \bar{\mathbf{a}}_0^{H} \mathbf{E}_N \mathbf{E}_N^{H} \bar{\mathbf{a}}_0. \tag{13}$$

According to constraint (13), convex optimization problem is constructed as

$$\begin{aligned} &\min_{\mathbf{e}_{\perp}} \left(\bar{\mathbf{a}}_0 + \mathbf{e}_{\perp}\right)^{\mathrm{H}} \mathbf{E}_N \mathbf{E}_N^{\mathrm{H}} \left(\bar{\mathbf{a}}_0 + \mathbf{e}_{\perp}\right), \\ &\text{s.t.}\, \bar{\mathbf{a}}_0^{\mathrm{H}} \mathbf{e}_{\perp} = 0 \\ &\left(\bar{\mathbf{a}}_0 + \mathbf{e}_{\perp}\right)^{\mathrm{H}} \mathbf{E}_N \mathbf{E}_N^{\mathrm{H}} \left(\bar{\mathbf{a}}_0 + \mathbf{e}_{\perp}\right) \le \bar{\mathbf{a}}_0^{\mathrm{H}} \mathbf{E}_N \mathbf{E}_N^{\mathrm{H}} \bar{\mathbf{a}}_{0.} \end{aligned} \tag{14}$$

The optimization problem (14) is a quadratic convex optimization problem, which can be solved with the openly accessed convex optimization toolbox CVX [22].

Then, the estimated steering vector can be obtained as

$$\hat{\mathbf{a}}_0 = \bar{\mathbf{a}}_0 + \mathbf{e}_{\perp}. \tag{15}$$

3.3. Interference Steering Vector Estimation. First, the angle and polarization parameters of the interferences need to be estimated to obtain the interference steering vectors. In array processing, high-resolution algorithm multiple signal classification (MUSIC) [23] is usually used to estimate source DOA. In the joint domain of space and polarization, the MUSIC joint spectrum can be expressed as

$$P(\theta, \phi, \gamma, \eta) = \frac{1}{\mathbf{a}_{sp}^{H}(\theta, \phi, \gamma, \eta) \mathbf{E}_N \mathbf{E}_N^{H} \mathbf{a}_{sp}(\theta, \phi, \gamma, \eta)}. \tag{16}$$

We assume that Θ_0 is neighbourhood of the desired signal location in the space-polarization domain:

$$\Theta_0 = \left\{\theta, \phi, \gamma, \eta | \hat{\theta}_0 - \delta_0 \le \theta \le \hat{\theta}_0 + \delta_0, \hat{\phi}_0 - \delta_0 \le \phi \le \hat{\phi}_0 + \delta_0, \hat{\gamma}_0 - \delta_0 \le \gamma \le \hat{\gamma}_0 + \delta_0, \hat{\eta}_0 - \delta_0 \le \hat{\eta} \le \hat{\eta}_0 + \delta_0\right\}, \tag{17}$$

where δ_0 is a preset value.

By performing search in joint domain beyond Θ_0, $M-1$ spectrum peaks are obtained. From the locations of the $M-1$ peaks, we can get the estimated interference parameter $(\hat{\theta}_i, \hat{\phi}_i, \hat{\gamma}_i, \hat{\eta}_i)$, $i = 1, 2, \cdots, M-1$. These estimations may have errors. In addition, steering vectors based on these estimations will lead to serious deterioration of the joint oblique projection method. So, the joint steering vectors need to be reestimated to enhance the robustness against estimation error.

Because the covariance matrix of the ith interference signal $\mathbf{R}_i$ cannot be obtained directly from the received mixed signal, it should be restructured to further calibrate the ith interference steering vector.

We define Θ_i to represent the neighbourhood of the ith interference centered on $(\hat{\theta}_i, \hat{\phi}_i, \hat{\gamma}_i, \hat{\eta}_i)$:

$$\Theta_i = \left\{ (\theta, \phi, \gamma, \eta) | \hat{\theta}_i - \delta \le \theta \le \hat{\theta}_i + \delta, \hat{\phi}_i - \delta \le \phi \le \hat{\phi}_i + \delta, \hat{\gamma}_i - \delta \le \gamma \le \hat{\gamma}_i + \delta, \hat{\eta}_i - \delta \le \hat{\eta} \le \eta_i + \delta \right\}, \tag{18}$$

where δ is a preset value larger than the estimated parameter error. So, the actual parameters of ith interference can be ensured to locate in Θ_i. Then, the estimated covariance matrix of ith interference $\hat{\mathbf{R}}_i$ is obtained by integrating in Θ_i:

$$\hat{\mathbf{R}}_i = \int\int\int\int_{(\theta,\phi,\gamma,\eta)\in\Theta_i} \mathbf{a}_{sp}(\theta, \phi, \gamma, \eta) \mathbf{a}_{sp}^{H}(\theta, \phi, \gamma, \eta) P(\theta, \phi, \gamma, \eta) \mathrm{d}\gamma\, \mathrm{d}\eta\, \mathrm{d}\theta\, \mathrm{d}\phi. \tag{19}$$

In practical, the integral operation represented by formula (19) is usually replaced by discrete summation in practical operation to get an approximate result:

$$\hat{\mathbf{R}}_i = \sum\sum\sum \sum_{(\theta,\phi,\gamma,\eta)\in\Theta_i} P(\theta, \phi, \gamma, \eta) \mathbf{a}_{sp}(\theta, \phi, \gamma, \eta) \mathbf{a}_{sp}^{H}(\theta, \phi, \gamma, \eta) \Delta\theta \Delta\phi \Delta\gamma \Delta\eta. \tag{20}$$

Since the covariance matrix only contains the ith interference, the joint steering vector of ith interference can be estimated by eigendecomposition of $\widehat{\mathbf{R}}_i$:

$$\widehat{\mathbf{R}}_i = \sum_{k=1}^{2N} \widehat{\lambda}_{ik} \widehat{\mathbf{v}}_{ik} \widehat{\mathbf{v}}_{ik}^{\mathrm{H}}, \tag{21}$$

where $\widehat{\lambda}_{ik}$ and $k = 1, 2, \cdots, 2N$ represent the eigenvalues in descending order and $\widehat{\mathbf{v}}_{ik}$ is the eigenvector corresponding to $\widehat{\lambda}_{ik}$.

According to the properties of subspace theory, the eigenvector $\widehat{\mathbf{v}}_{i1}$ corresponding to the largest eigenvalue coincides with the steering vector of ith interference. Then, the joint steering vector of each interference can be estimated one by one.

3.4. Optimal Weight Calculation. Through the above-mentioned process, the steering vector of desired signal has been obtained. The projection subspace can be defined as $\mathbf{S} = \widehat{\mathbf{a}}_0$. The interference subspace can be defined as $\mathbf{V} = [\widehat{\mathbf{v}}_{11}, \widehat{\mathbf{v}}_{21}, \cdots, \widehat{\mathbf{v}}_{M1}]$. So, the joint oblique projection operator $\mathbf{Q}_{SV}$ can be obtained by equation (9) and (10).

Then, weight vector $\mathbf{w}$ can be calculated as

$$\mathbf{w} = \left(\left(\widehat{\mathbf{a}}_0^H \widehat{\mathbf{a}}_0 \right)^{-1} \widehat{\mathbf{a}}_0^H \mathbf{Q}_{SV} \right)^H. \tag{22}$$

The proposed method can be summarized as follows:

(1) Calculate $\widehat{\mathbf{R}}$ by (8) and take eigendecomposition of it to obtain the noise subspace $\mathbf{E}_N$

(2) Solve problem (14) to modify the steering vector of desired signal

(3) Estimate the *interference parameters* by the MUSIC spectrum in (16)

(4) Reconstruct the covariance matrix of ith interference $\widehat{\mathbf{R}}_i$ by (20) and take eigendecomposition of it to obtain the estimation of steering vector of the ith interference

(5) Construct the oblique projection operator $\mathbf{Q}_{SV}$ by using the estimated steering vectors and obtain the weight vector $\mathbf{w}$ of the beamformer by (22)

4. Simulation Results

In this section, numerical simulations are performed to analyse the effectiveness of the proposed method. MEPS in [13] and methods in [17, 18] are only processed in the spatial domain, and they are extended to the space-polarization joint domain as comparison methods. The methods CMRB in [12] and SPMVDR are also performed as comparison methods.

We consider a semicircular conformal array located in the XOY plane as shown in Figure 2. The number of array elements is 20, the distance between adjacent array elements is 0.5 λ, and the radius of the circular array is about 1.59 λ. The lowest-order circular patch antenna is selected as the antenna element model of the semicircular conformal array. The polarization components of the antenna radiating (θ, ϕ) direction are

$$\begin{cases} g_\theta(\theta, \phi) = \left[J_2(\pi d \sin\ \theta/\lambda) - J_0(\pi d \sin\ \theta/\lambda) \right] [\cos\phi - j \sin\phi], 0 \le \theta \le \pi/2, \\ g_\phi(\theta, \phi) = \left[(\pi d \sin\ \theta/\lambda) + J_0(\pi d \sin\ \theta/\lambda) \right] \cos\theta [\sin\phi - j \cos\phi], 0 \le \theta \le \pi/2, \\ g_\theta(\theta, \phi) = g_\phi(\theta, \phi) = 0, \theta > \pi/2. \end{cases} \tag{23}$$

where J_0 and J_2 are the zero-order and second-order Bessel functions of the first kind, respectively.

The beam pattern function of dual-polarized conformal array is a four-dimensional function of the parameter $(\theta, \phi, \gamma, \eta)$. To simplify the calculation, the parameters (θ, η) of all the signals are fixed as $(80^\circ, 30^\circ)$. We assume the space-polarization information of the mainlobe interference and side-lobe interference are $(\phi_1, \gamma_1) = (21^\circ, 70^\circ)$ and $(\phi_2, \gamma_2) = (60^\circ, 60^\circ)$, the interference-to-noise ratios (INR) are both 30 dB. The integral range δ is set to 5°.

4.1. Mismatch due to Direction and Polarization Parameter Error. We set the prior space-polarization information of the desired signal as $(\phi_0, \gamma_0) = (20^\circ, 10^\circ)$ and the actual information as $(\phi_0, \gamma_0) = (23^\circ, 5^\circ)$. The SNR is 10 dB. The number of snapshots is 500. The beam pattern of the proposed method is shown in Figure 3. Obviously, the proposed method forms deep null in the correct position of each interference. The gain of desired signals is guaranteed.

The spatial domain error of desired signal is assumed to be uniformly distributed over $[-4^\circ, 4^\circ]$, and the polarized domain error is uniformly distributed over $[-8^\circ, 8^\circ]$. The number of Monte Carlo experiments is 200. Figure 4 shows the output SINR of each method versus input SNR varies from -5 dB to 25 dB in the condition of the number of snapshots $L = 500$. Figure 5 shows that the output SINR of each method versus the number of snapshots varies from 50 to 1000, while SNR = 10 dB. It can be seen that the SINR of the proposed method is still better than others. The performance of the proposed method does not decrease as SNR increases. In addition, the proposed method still has the best performance with any number of snapshots.

4.2. Mismatch due to Steering Vector Random Error. This part analyses the performance in the case of mismatch caused by the steering vector random error. It can be considered as a general type of mismatch in practice. Consider generating actual steering vector by adding a random error vector to the assumed steering vector $\mathbf{a}_0 = \overline{\mathbf{a}}_0 + \mathbf{e}$, where $\mathbf{e}$ is the random

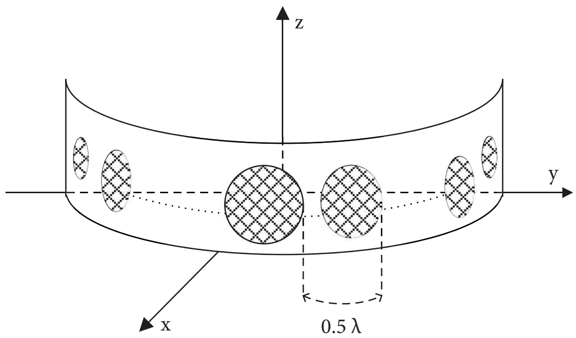

Figure 2: Dual-polarized semicircular patch antenna array.

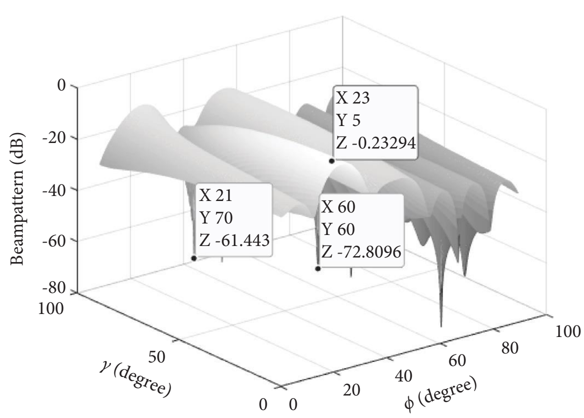

Figure 3: The polarization-space beampattern of the proposed method with a parameter error of the desired signal.

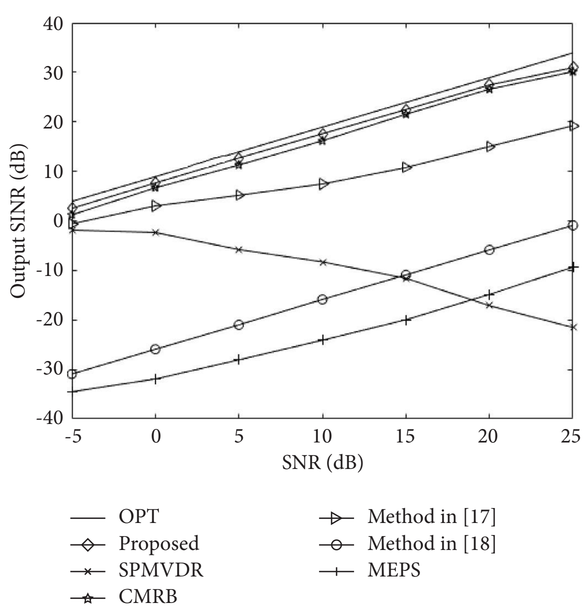

Figure 4: Output SINR versus SNR with a parameter error of the desired signal.

error vector. The random error $\mathbf{e}$ can be expressed as $\mathbf{e} = \epsilon/\sqrt{2N}\,[e^{j\psi_1}, e^{j\psi_2}, \cdots, e^{j\psi_{2N}}]^T$, where ϵ follows a uniform distribution in $[0, 2]$ and phases ψ_k and $k = 1, \cdots, 2N$ are uniformly distributed in $[0, 2\pi]$. They are independent of

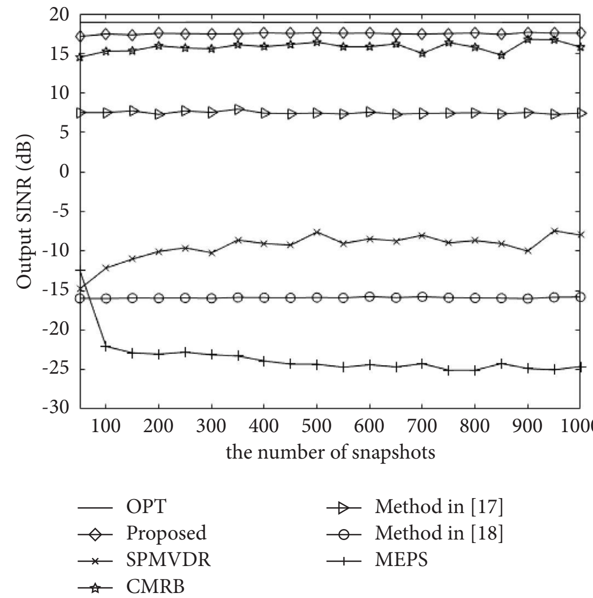

Figure 5: Output SINR versus the number of snapshots with a parameter error of the desired signal.

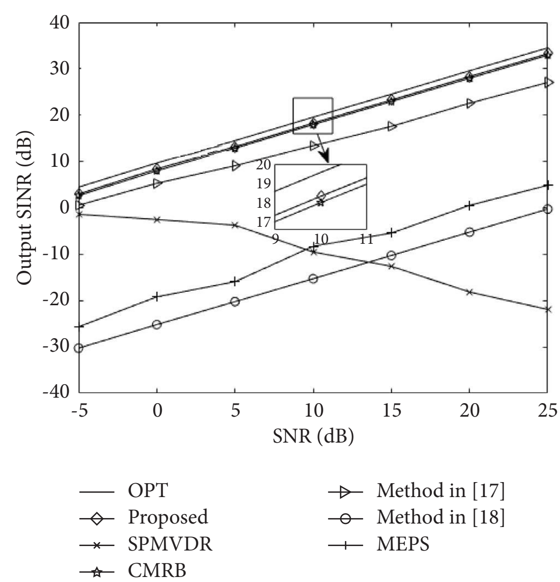

Figure 6: Output SINR versus SNR with the steering vector random error.

each other. The number of Monte Carlo experiments is 200. Figure 6 shows that the output SINR of each method versus input SNR varies from -5 dB to 25 dB with $L = 500$. Figure 7 shows that the output SINR of each method versus the number of snapshots varies from 50 to 1000, while SNR = 10 dB. Obviously, the performance of the proposed method is still better than other methods in this scenario.

Since MEPS and methods in [17, 18] are proposed based on ULA in the spatial domain, they deteriorate in varying degrees when they are extended to the space-polarization joint domain. For MEPS, although it can suppress interferences in joint domains, the noise power generated by MEPS is much larger than the signal power leading to

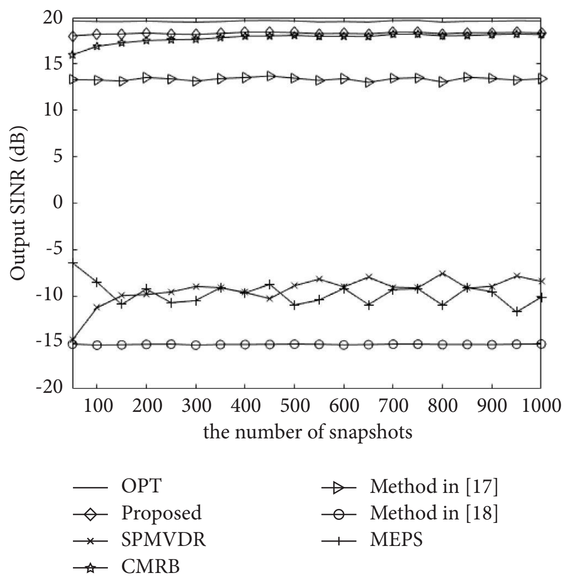

Figure 7: Output SINR versus the number of snapshots with the steering vector random error.

significant performance degradation. Due to resolution of polarization domain being lower than spatial domain, the method in [18] for estimating steering vector of desired signal lost its effectiveness in joint domains. The performance degradation of the method in [17] is not serious, as it is not targeted at ULA and can be applied in various situations. Compared to other beamformers, the proposed beamformer can provide higher output SINR and approach the optimization result in various situations.

5. Conclusions

Aiming at the deficiencies of current mainlobe interference suppression methods, a novel beamforming method is proposed in the space-polarization joint domain base on dual-polarized conformal array. To improve the robustness of oblique projection filtering, desired signal and interference steering vectors are estimated to establish joint oblique projection operator. For desired signal mismatch, a convex optimization problem is constructed to make estimated steering vector close to actual steering vector. For reducing the impact of the estimated error of the interference parameters, the interference covariance matrix is reconstructed by integrating in the neighborhood of estimated parameter of interference to further estimate the interference steering vector. Then, they are used to construct joint oblique projection filter to calculate the optimal weight vector. Simulation experiments show that the proposed method can effectively suppress the mainlobe and side-lobe interferences. It is robust to the errors of interference parameters and the mismatch of steering vector of desired signal. However, the proposed method needs four-dimensional search and integration in general situations, which brings huge computation in practical applications. How to deal with this issue should be considered in the future.

Data Availability

The image data used to support the findings of this study are available from the corresponding author upon request.

Conflicts of Interest

The authors declare that they have no conflicts of interest.

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