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# Cascaded Interference and Multipath Suppression Method Using Array Antenna for GNSS Receiver

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**ABSTRACT** The space-time adaptive processing method has been widely used for both narrowband and wideband interference suppression in GNSS receiver equipped with array antenna. However, the use of space-time filter may cause satellite signal distortion and lead to bias in pseudo-range measurement. In addition, multipath is one of the main error sources that will have a serious influence on the performance of the GNSS receiver. For precise applications, the error induced by space-time filter and multipath should be reduced. In this paper, a cascaded interference and multipath suppression method using an array antenna are proposed. In the first stage, the interference is suppressed by projecting the received array signal into the interference orthogonal subspace using subspace projection method. In the second stage, the directions of arrival (DOA) of the direct signal and the multipath signal are estimated using sparse recovery method. The DOA parameters enable the receiver to construct a space-time distortionless beamformer, which is used to null the multipath signals and maximize the signal-to-noise ratio. The method does not require the prior knowledge of the direction of the direct signal and achieves unbiased measurement while suppressing the interference and multipath. The simulation results demonstrate that the proposed method can achieve the interference and multipath suppression effectively, and have a distortionless response for the direct signal.

**INDEX TERMS** GNSS receiver, interference suppression, multipath suppression, array antenna, sparse recovery.

## I. INTRODUCTION

Global Navigation Satellite System (GNSS) has been widely used for high precision positioning, navigation, timing, and other applications. The satellite signal received by GNSS receiver is very weak (typically 20dB below the noise level), which is susceptible to intentional and unintentional interference.

The interference and the multipath are two of the main error sources, which will downgrade the performance in GNSS receiver [1]. In some high precision application, e.g. high precision monitoring receivers and precise positioning applications, accurate and unbiased pseudo-range measurement are required. Therefore the bias caused by the interference and multipath need to be mitigated [2].

Adaptive array processing is a promising method and has been used for interference suppression in GNSS receiver, which can mitigate many more narrowband and wideband interference [2]. By adding finite impulse filter (FIR) after each antenna, it leads to the adaptive space-time processing (STAP) technique, whose degrees of freedom (DoF) is increased compared with spatial-only filtering [3]. However, the satellite signal may be distorted by using space-time filters in each channel, which will induce bias into pseudo-range measurement [4]. The beamforming methods based on minimum variance principle are most widely used in GNSS receiver, which has minimized the output power. Among them, the minimum variance distortion response (MVDR) method applies a linear constraint on the weight to steer the gain in the satellite signal direction, and the power inversion (PI) method take a certain antenna as the output. PI is a blind beamforming method, in which the prior knowledge of the directions of satellite signals is not needed.

In order to mitigate the bias induced by space-time filtering, many approaches are proposed [5]–[7]. In [6], an improved MVDR method is proposed by constraining the tap number to be odd and taking the middle tap as the output of the array, which can achieve linear response of the space-time filter. However, the parameters of the direction of the direct signal and the array manifold parameters are often

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unavailable in practice, which has limited its application. In [8], a distortionless blind beamforming method is proposed by constraining its coefficients to ensure linearity of the space-time processor response, which requires half of the DoFs. In [9], a distortionless space-time process method is proposed by employing a cascaded filter following the STAP filter, whose frequency response is the conjugate of that of the STAP filter. The drawback of using blind beamforming method is that it may lead to satellite signal energy loss. However, all above methods cannot suppress the multipath error efficiently.

The multipath error can be up to dozens of meters in the pseudo-range measurements, which is one of the main error sources that affect positioning accuracy in GNSS applications. Lots of multipath methods based on single antenna have been proposed, e.g., narrow correlator technology [10], strobe correlator [11], multipath likelihood Delay-Lock Loop (MEDLL) [12], and so on. The drawback of the multipath methods based on single antenna is that the short delay (<0.1chip) multipath signal cannot be suppressed [1]. Array antenna for multipath suppression has been studied in many literatures [13]–[15]. However, most of these methods only suppress the multipath without considering the presence of the interference. A two-stage beamformer for interference and multipath mitigation method is proposed in [15], a modified MVDR beamformer employing overlapping sub-arrays is used to mitigate the multipath components after interference suppression stage. There are two drawbacks for this method, the prior knowledge of the direct signal direction is required, and it is only suitable for space-only process, which cannot reduce the bias induced by the space-time processer. From the above analysis, distortionless interference suppression and multipath suppression methods still require further research.

This paper proposes a cascaded interference and multipath suppression method using array antenna. In the first stage, the interference is suppressed using subspace projection method, and the eigenvalue decomposition process is replaced by the inverse matrix of the covariance matrix to reduce the computational complexity. In the second stage, the array signals after interference suppression are firstly correlated with the local replica signal. Then sparse recovery method is used for direction of arrival (DOA) estimation. There are only limited numbers of incident directions for multipath signal, which is sparse in space and can be reconstructed by sparse recovery method [16]. After that, a space-time beamformer is constructed to steer the array beam towards the direct signal and null towards the multipath signals. The proposed approach has three advantages compared to the traditional methods:

(1) The interference and multipath can be suppressed without using the prior knowledge of the direction of the desired signal and the array manifold parameters.

(2) A distortionless space-time beamformer is constructed using the steer vector estimations of the direct signal and multipath, which can put null in the direction of the multipath and achieve distortionless response of the desired signal. The bias induced by space-time processer can be eliminated.

(3) The direct signal is enhanced and can achieve maximum signal-to-noise ratio compared with other blind beamforming method.

This paper is organized as follows. In Section II, the models of multipath signal and array signal are presented. The interference suppression method based on subspace projection is described in Section III. The multipath suppression method based on sparse recovery method is analyzed in Section IV. In Section V, the simulation results are given to verify the effectiveness of the proposed method. Section VI concludes this paper.

Throughout this paper, the following notations are adopted: $(\cdot)^H$, $(\cdot)^T$, $E[\cdot]$, and $\odot$ denote the operations of complex conjugate transpose, transpose, statistical expectation, and Hadamard product, respectively. The bold lowercase letters stand for vectors and capital bold letters stand for matrices.

## II. SIGNAL MODEL

Without loss of generality, only one satellite signal is considered. Suppose there are one direct signal, $M$ multipath signals, and $Q$ jammers incident on an uniform linear array (ULA) with $N$ elements from the far field, the compounded signal received by GNSS receiver can be written as

$$\mathbf{x}(t) = \sum_{m=0}^{M} \alpha_m \mathbf{a}(\theta_m) s(t-\tau_m) + \sum_{q=1}^{Q} \mathbf{a}(\theta_q) j_q(t) + \mathbf{n}(t) \tag{1}$$

where $\mathbf{x}(t) = [x_1(t), x_2(t), \cdots, x_N(t)]^T$ is the array data vector with size of $N \times 1$. $s(t)$ is the model of GNSS signal, $\tau_m$ $(m = 0, 1, \cdots, M)$ and $\alpha_m$ $(\alpha_m \leq 1)$ are the time delay and the correlation coefficient of the $m$th signal component, respectively. The subscript zero stands for the direct signal. $j_q(t)$ is the $q$th interference component. $\mathbf{n}(t)$ represents the additional white Gaussian noise vector. $\mathbf{a}(\theta)$ is the steering vector, which can be expressed as

$$\mathbf{a}(\theta) = \left[1, e^{-j2\pi d \sin(\theta)/\lambda}, \cdots, e^{-j2\pi d(N-1)\sin(\theta)/\lambda}\right]^T \tag{2}$$

where $d$ is the array element spacing, $\lambda$ is the wavelength of the signal, $\theta$ is the angle of arrival.

The GNSS signal can be modeled as

$$s(t-\tau) = A \cdot D(t-\tau) c(t-\tau) e^{j(2\pi f_d t + \varphi)} \tag{3}$$

where $D(\cdot)$ is the navigation bit. In this paper, the correlation process is realized within one data bit, therefore, the navigation bit is considered as a constant equal to 1. $c(\cdot)$ is the pseudo-random noise (PRN) code. $A, f_d$, and $\varphi$ are the signal amplitude, Doppler frequency, and carrier phase of the signal, respectively.

For STAP method with $L$ taps for time filter, the weight coefficients with $NL$ elements can be expressed as

$$\mathbf{w} = [w_{1,1}, w_{1,2}, \cdots, w_{1,N}, w_{2,1}, \cdots, w_{L,N}]^T \tag{4}$$

For each snapshot, the received signal vector contains $NL$ elements, which can be written as

$$\tilde{\mathbf{x}} = [\mathbf{x}^T(t), \mathbf{x}^T(t-T_s), \cdots, \mathbf{x}^T(t-(L-1)T_s)]^T \tag{5}$$

where $T_s$ denotes the sampling duration.

The output of the array after weighting is

$$y(t) = \mathbf{w}^H \tilde{\mathbf{x}} \tag{6}$$

## III. INTERFERENCE SUPPRESSION

The subspace projection method is one of the blind beamforming methods. The principle of the method is to construct the interference subspace by performing eigenvalue decomposition on the covariance matrix of the array signal and project the array signal into the orthogonal subspace of the interference. The array signal covariance matrix can be expressed as

$$\mathbf{R} = E\left[\tilde{\mathbf{x}}\tilde{\mathbf{x}}^H\right] \tag{7}$$

Generally, the navigation signal received by the receiver is very weak and submerged under noise, which can be ignored compared with the interference and noise. Therefore, the array covariance matrix is mainly determined by interference and noise. Assuming that the interference is independent of each other and between the interference and the noise, the array signal covariance matrix is approximately equal to the interference covariance matrix and the noise covariance matrix.

$$\begin{aligned}\mathbf{R} &= \sum_{q=1}^{Q}(P_q - \sigma_n^2)\mathbf{u}_q\mathbf{u}_q^H + \sigma_n^2\mathbf{I}_{NL\times NL}\\ &= \mathbf{U}_I \sum \mathbf{U}_I^H + \sigma_n^2\mathbf{I}_{NL\times NL}\end{aligned} \tag{8}$$

where $\mathbf{U}_I = [\mathbf{u}_1, \mathbf{u}_2, \cdots, \mathbf{u}_Q]$, $\mathbf{u}_q$ is the eigenvector of the interference, $\sum = diag\left(P_1 - \sigma_n^2, P_2 - \sigma_n^2, \cdots, P_Q - \sigma_n^2\right)$ is the diagonal matrix with the power difference between interference and noise along its diagonal, $\sigma_n^2$ denotes the noise power, and $\mathbf{I}_{NL\times NL}$ stands for the identity matrix.

If $P_q \gg \sigma_n^2, q = 1, \cdots, Q$, the inverse of the covariance matrix can be expressed as [17]:

$$\mathbf{R}^{-1} \approx \frac{1}{\sigma_n^2}\left(\mathbf{I}_{NL\times NL} - \mathbf{U}_I(\mathbf{U}_I^H\mathbf{U}_I)^{-1}\mathbf{U}_I^H\right) \tag{9}$$

From (9), it can be found that the inverse of the covariance matric can be approximated as the interference null space. Therefore, the eigenvalue decomposition process can be replaced by the inverse matrix of the covariance matrix to reduce the computational complexity. The interference can be suppressed by projecting the received signal onto the interference null space.

The array signal after interference suppression can be expressed as

$$\begin{aligned}\hat{\mathbf{x}} &= \mathbf{R}^{-1}\tilde{\mathbf{x}}\\ &\approx \begin{bmatrix}\mathbf{R}_{1,1}\mathbf{x}(t)\\ \mathbf{R}_{2,2}\mathbf{x}(t-T_s)\\ \vdots\\ \mathbf{R}_{L,L}\mathbf{x}\left(t-(L-1)\cdot T_s\right)\end{bmatrix}\\ &= \sum_{m=0}^{M}\alpha_m\begin{bmatrix}\mathbf{R}_{1,1}\mathbf{a}(\theta_m)s\left(t-\tau_m\right)\\ \mathbf{R}_{2,2}\mathbf{a}(\theta_m)s\left(t-\tau_m-T_s\right)\\ \vdots\\ \mathbf{R}_{L,L}\mathbf{a}(\theta_m)s\left(t-\tau_m-(L-1)\cdot T_s\right)\end{bmatrix}\end{aligned} \tag{10}$$

where

$$\mathbf{R}^{-1} = \begin{bmatrix}\mathbf{R}_{1,1} & \mathbf{R}_{1,2} & \cdots & \mathbf{R}_{1,L}\\ \mathbf{R}_{2,1} & \mathbf{R}_{2,2} & \cdots & \mathbf{R}_{2,L}\\ \vdots & \vdots & \ddots & \vdots\\ \mathbf{R}_{L,1} & \mathbf{R}_{L,2} & \cdots & \mathbf{R}_{L,L}\end{bmatrix}.$$

The inverse of the covariance matrix can be considered as the noise space. According to the noise characteristics, the noise energy is mainly distributed on the diagonal matrix, and the values of other sub-matrices are approximately to be zeros. The interference is considered to be suppressed completely, and the influence of the noise term is omitted for the sake of simplicity.

## IV. MULTIPATH SUPPRESSION

After interference suppression, the dimension of the array signal is not changed, which can be used for generating additional beamformer for multipath suppression.

### A. DOA ESTIMATION BASED ON SPARSE RECOVERY METHOD

The local replica signal can be modeled as

$$s_L(t-\hat{\tau}) = c(t-\hat{\tau})e^{j(2\pi\hat{f}_d t+\hat{\varphi})} \tag{11}$$

where $\hat{\tau}$, $\hat{f}_d$, and $\hat{\varphi}$ denote the estimation values of the code delay, Doppler frequency, and carrier phase, respectively.

The local replica is then multiplied by the received signal, and the correlation results of the array can be modeled as

$$\begin{aligned}\mathbf{r} &= E\left[\hat{\mathbf{x}}(t)\odot\mathbf{s}_L^H\right]\\ &= \sum_{m=0}^{M}\alpha_m\begin{bmatrix}\mathbf{R}_{1,1}\mathbf{a}(\theta_m)r(\Delta\tau_m)\\ \mathbf{R}_{2,2}\mathbf{a}(\theta_m)r(\Delta\tau_m)\\ \vdots\\ \mathbf{R}_{L,L}\mathbf{a}(\theta_m)r(\Delta\tau_m)\end{bmatrix}\\ &\approx \sum_{m=0}^{M}\alpha_m\mathbf{R}^{-1}\mathbf{a}_{stap}(\theta_m)r(\Delta\tau_m)\end{aligned} \tag{12}$$

where $\Delta\tau_m = \hat{\tau} - \tau_m$ denotes the code phase estimation error of the $m$th component, $r(\cdot)$ denotes the auto-correlation function (ACF) of the navigation signal, $\mathbf{a}_{stap}(\theta_m) = [\underbrace{\mathbf{a}^T(\theta_m), \cdots, \mathbf{a}^T(\theta_m)}_{L}]^T$ is the steer vector of space-time adaptive processor, $\mathbf{s}_L = [\mathbf{I}_{1\times N}\cdot s_L(t-\hat{\tau}), \cdots, \mathbf{I}_{1\times N}\cdot s_L(t-\hat{\tau}-(L-1)T_s)]$. The carrier is considered to be completely wiped-off.

The weighted output of (12) can be expressed as

$$y_{out} = \mathbf{w}^H \mathbf{r} = \sum_{m=0}^{M} \alpha_m \mathbf{w}^H \bar{\mathbf{a}}(\theta_m) r(\Delta\tau_m) \tag{13}$$

where $\bar{\mathbf{a}}(\theta_m) = \mathbf{R}^{-1}\mathbf{a}_{stap}(\theta_m)$. If $\mathbf{w} = \bar{\mathbf{a}}(\theta_0)$, the correlation value of the direct signal is maximized. Similarly, if $\mathbf{w} = \bar{\mathbf{a}}(\theta_m)$, the correlation value of the $m$th multipath is maximized, while, the other signals are suppressed. Therefore, the steering vector can be used as array weight coefficient for signal detection.

As for the ULA, the incident angle of the incoming signal ranges from -90 degree to 90 degree. If the space is divided into $K$ discrete direction sets, the spatial angle intervals is $\Delta\theta = 180/K$. The redundancy matrix of the direction steering vector can be expressed as

$$\mathbf{A} = [\bar{\mathbf{a}}(\theta_1), \bar{\mathbf{a}}(\theta_2) \cdots, \bar{\mathbf{a}}(\theta_K)] \tag{14}$$

where $\theta_k = -90 + (k-1) \cdot \Delta\theta, k = 1, 2, \cdots, K$.

The matrix $\mathbf{A}$ contains the steering vectors of all the discrete angle direction. The correlation vector $\mathbf{r}$ can be reconstructed using $\mathbf{A}$ as

$$\mathbf{r} \approx \mathbf{A} \cdot \boldsymbol{\alpha} \tag{15}$$

where $\boldsymbol{\alpha}$ denotes the coefficient vector of size $K \times 1$, there are only $M+1$ non-zero elements, which represent the direct and the multipath components. The amplitudes of the nonzero elements represent the correlation value weighted by the corresponding column steering vector in matrix $\mathbf{A}$.

The coefficient vector $\boldsymbol{\alpha}$ can be recovered by solving the following $\ell_0$-minimization problem [18]:

$$\hat{\boldsymbol{\alpha}} = \arg\min_{\boldsymbol{\alpha}} ||\boldsymbol{\alpha}||_0 \quad s.t. \; ||\mathbf{r} - \mathbf{A}\boldsymbol{\alpha}||_2 \leq \sigma \tag{16}$$

where $\sigma$ is the standard deviation of the noise. $\ell_p$-norm of the vector $\boldsymbol{\alpha}$ is defined as $||\boldsymbol{\alpha}||_p = \sqrt[p]{\sum_i |\alpha_i|^p}$, the $\ell_0$-norm is simply defined as the number of nonzero (significant) elements.

It is not easy to solve (16), while the $\ell_0$-minimization problem can be converted to solving the $\ell_1$-minimization problem [19]:

$$\hat{\boldsymbol{\alpha}} = \arg\min_{\boldsymbol{\alpha}} ||\boldsymbol{\alpha}||_1 \quad s.t. \; ||\mathbf{r} - \mathbf{A}\boldsymbol{\alpha}||_2 \leq \sigma \tag{17}$$

The above optimization can be solved by convex optimization method or greedy algorithm. The OMP method [20] is used in this paper for solving (17), which is one of the greedy algorithms and can achieve higher computational efficiency than other sparse recovery methods. The principle of OMP is selecting the atom that best matches the observed signal from the redundant matrix by iterative method.

### B. PERFORMANCE ANALYSIS

#### 1) ESTIMATION ACCURACY

The received signal energy, the number of the array elements, and the correlation time are the three parameters that have important impact on the estimation accuracy of DOA. In order to evaluate the impact of these parameters on the estimation performance, Monte Carlo simulation is presented using 500 independent trails, and the proposed method is compared with spatial smoothing MUSIC method [21], which can achieve the DOA estimation of coherent signal using spatial smoothing method for de-correlation. In the simulation, there are one direct signal and one multipath, and the incident angles are 0 degree and 30 degree, respectively. The energy of the multipath is 3dB lower than the direct signal. A ULA with half-wavelength spacing is used. The root-mean-square error (RMSE) of the DOA estimation results versus carrier-to-noise ratio($C/N_0$) of the direct signal are shown in Fig. 1. From Fig. 1, we can observe that estimation accuracy will be increased along with the increasing number of antenna elements and the $C/N_0$. In the same simulation conditions, the performance of OMP is better than the MUSIC method. When using the array with 12 elements, the estimation error is less than 1 degree even in low $C/N_0$, which satisfies the requirements of the algorithm proposed in this paper. Noting that $C/N_0$ is defined as the ratio of the power of the signal $C$ to the noise power $N_0$ in a 1Hz bandwidth.

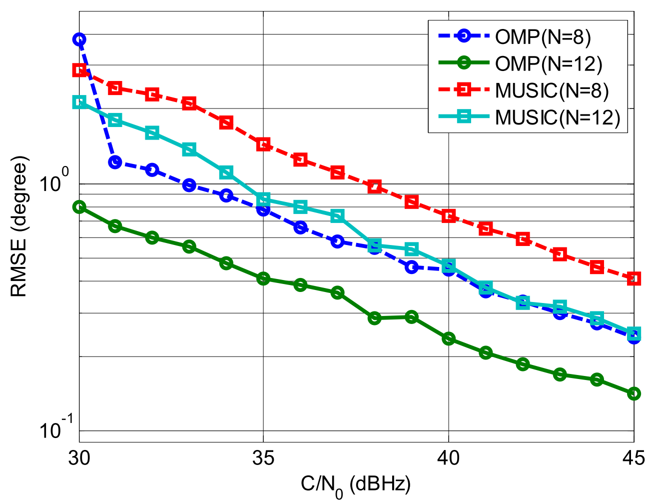


**FIGURE 1.** RMSE results of DOA estimation versus $C/N_0$ with different array size.

#### 2) ALGORITHM COMPLEXITY

DOA Estimation using OMP method is the most complicated part of the proposed method. When building the redundant matrix, each of the elements in the matrix is multiplied by the inverse matrix of the covariance matrix, which needs $KN^2L^2$ complex multiplication and $KNL(NL-1)$ complex addition operations. The number of iterations used in OMP method is $M+1$. In each iteration period, it requires $K$ times vector multiplications, and each contains $NL$ times complex multiplication and $NL-1$ times complex addition operations. Therefore, the total computational amount is $K(M+NL+1)NL$ times multiplication operations and $K(M+NL+1)(NL-1)$ complex addition operations. Since $M < N \ll K$, the total computation time complexity is about $O(K(M+NL+1)NL)$, which mainly depends on the spatial direction interval. The smaller the spatial angle interval, the higher the estimation accuracy of DOA and the larger the complexity of the algorithm.

### C. DIRECT SIGNAL DETECTION

As for noise-free navigation signal, the largest value in $\hat{\boldsymbol{\alpha}}$ represents the direct signal, and other nonzero elements represent the multipath components, which means the amplitude estimation values can be used for direct signal detection and separation from multipath components. However, the influence of the noise cannot be ignored. Fig. 2 shows the Monte Carlo simulation results. The simulation parameters are the same with that in Section IV (B). Fig 2(a) shows the RMSE of the amplitude estimation result of the weighted correlation value, and Fig. 2(b) shows the detection probability of the direct signal. From Fig. 2(a), it can be found that the proposed method can achieve precise amplitude estimation. The estimation accuracy will increase along with the increase of the number of the array elements and $C/N_0$. Fig. 2(b) shows that it has a high probability for successfully separating the direct signal and multipath. The detection probability is almost 100% when the $C/N_0$ is larger than 35dBHz and the array element number is larger than 8. When the $C/N_0$ or the number of array elements decreases, the detection performance decreases.

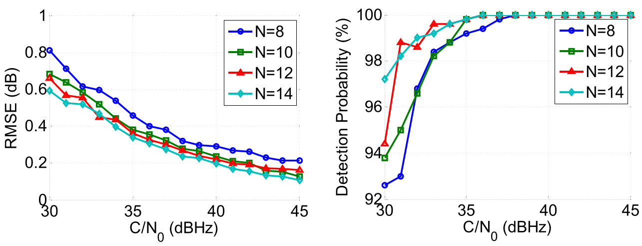


**FIGURE 2. Simulation results: (a) the RMSE of the amplitude estimation results; (b) detection probability of the direct signal.**

In order to improve the detection performance in low $C/N_0$, an additional strategy is used for judging the direct signal. In practice, the incident angle of the direct signal changes very slowly, which can be considered as a constant in a short time interval, e.g. a few milliseconds. Therefore, the estimated result of the incident angle is not used directly, but compared with the mean value of the past several results. Suppose the current estimation result is $\hat{\theta}_0$, and the past $P$ times estimation results are $\hat{\theta}_p, p = 1, \cdots, P$, the estimation error can be calculated as

$$\Delta = \hat{\theta}_0 - \frac{1}{P}\sum_{p=1}^{P}\hat{\theta}_p \tag{18}$$

Then $\Delta$ is compared with a threshold $Th$. If $\Delta < Th$, then $\hat{\theta}_0$ is used as the accurate estimation of the incident angle of the direct signal. Otherwise, if $\Delta \geq Th$, the last estimation results is used as the correct result instead of the current estimation value. $P$ is choose as 10 in this paper. The half-power beam width (HPBW) for a ULA with half-wavelength spacing is approximately $102/N$ degree [22], so the threshold $Th$ is defined as $51/N$ degree, which is half of HPBW. As for an array with 12 elements, the threshold is about 4.25 degree.

### D. BEAMFORMING METHOD USING ARRAY ANTENNA

In order to enhance the direct signal and null the multipath signal, the weight vector needs to satisfy the following expression.

$$\hat{\mathbf{w}}^H\bar{\mathbf{A}} = \mathbf{b} \tag{19}$$

where $\bar{\mathbf{a}} = [\mathbf{a}(\hat{\theta}_0), \mathbf{a}(\hat{\theta}_1), \cdots, \mathbf{a}(\hat{\theta}_M)]$, $\mathbf{a}(\hat{\theta}_0)$ is the steering vector estimation of the direct signal, and the rest are the steering vectors estimation results of the multipath signals. By constraining $\mathbf{b} = [1, 0, \cdots, 0]^T$, which means the output of direct signal is straight-through, but without the multipath signals, the solution of (19) can be expressed as

$$\hat{\mathbf{w}} = \bar{\mathbf{A}} \cdot (\bar{\mathbf{A}}^H \cdot \bar{\mathbf{A}})^{-1} \cdot \mathbf{b} \tag{20}$$

Using the same constraint in [6], i.e. constraining the tap number $L$ to be odd and taking the middle tap as the output of the array, the weight vector for space-time processing can be expressed as

$$\mathbf{w} = \left[\mathbf{0}_{1\times N(L-1)/2}, \hat{\mathbf{w}}^T, \mathbf{0}_{1\times N(L-1)/2}\right]^T \tag{21}$$

The output signal of the array weighted by (21) can be written as

$$y(t) = \mathbf{w}^H\mathbf{R}^{-1}\tilde{\mathbf{x}} \tag{22}$$

Taking (10) into (22), and supposing the multipath components are suppressed completely, the output of the array can be rewritten as

$$\begin{aligned} y(t) &\approx \hat{\mathbf{w}}^H\mathbf{R}_{(L-1)/2,(L-1)/2}\mathbf{x}\left(t-(L-1)T_s/2\right) \\ &= \mu \cdot s\left(t-\tau-(L-1)T_s/2\right) \end{aligned} \tag{23}$$

where $\mu = \hat{\mathbf{w}}^H\mathbf{R}_{(L-1)/2,(L-1)/2}\mathbf{a}(\theta_0)$ is a constant value. If there are no multipath signals, then $\mu = \mathbf{a}(\hat{\theta}_0)^H\mathbf{R}_{(L-1)/2,(L-1)/2}\mathbf{a}(\theta_0)/N$, which has a similar expression with the MVDR method [23]. Using the proposed method, the linear response of the space-time filter is achieved.

From (23), it can be found that after the proposed cascaded interference and multipath suppression method, the interference and the multipath can be suppressed; and at the same time, the direct signal can be processed without distortion and there has zero digital beaforming induced biases. It should be noted that the proposed method will introduce a constant pseudo-range bias $(L-1)T_s/2$, which can be calculated accurately.

If $L = 1$, the space-time processor becomes spatial only processor, which will not cause signal distortion. And the proposed method can still be used for multipath suppression and enhancing the direct signal.

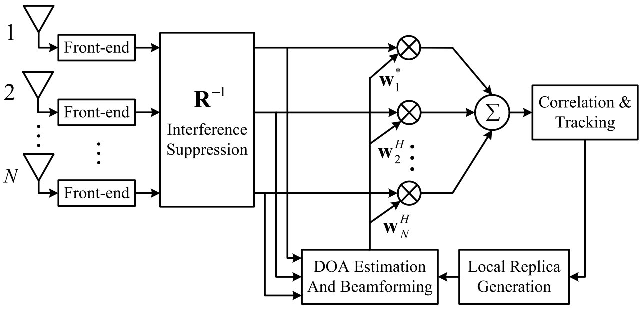


**FIGURE 3.** Block diagram of the proposed method.

### E. THE PROCEDURE OF THE PROPOSED METHOD

The diagram of the proposed algorithm is shown in Fig. 3 with the following steps.

1). Suppress the interferences using the inverse matrix of the covariance matrix of the array signal in (7). In practice, $\mathbf{R}$ is usually calculated by time averaging

$$\mathbf{R} \approx \frac{1}{I}\sum_{i=1}^{I}\mathbf{X}_i\mathbf{X}_i^H \tag{24}$$

where $I$ is the correlation times.

2). Generate the over-complete dictionary matrix $\mathbf{A}$ in (14).

3). Construct the local replica signal $s_L(t)$ in (11) using estimation of code delay $\hat{\tau}$, Doppler frequency $\hat{f}_d$, and carrier phase $\hat{\varphi}$ from the output of tracking loop.

4). Compute the correlation value vector in (12) by multiplying the replica with the received array signal after interference suppression.

5). Estimate the DOA of the direct signal and multipath signal using OMP method by solving (17).

6). Calculate the weight coefficient vector $\hat{\mathbf{w}}$ in (21).

7). Weight the array in (22) and sent the output into the correlation and tracking module.

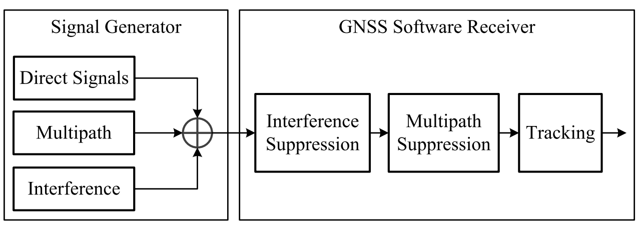


**FIGURE 4.** Flowchart of the simulation model.

## V. SIMULATION AND DISSCUSSION

The simulation modules are shown in Fig. 4. Firstly, the direct satellite signal, the multipath, and the interference are generated by a software signal generator, respectively. Then, the compounded signal is processed by the GNSS software receiver for interference and multipath suppression and signal tracking. The navigation signal used in this simulation is BeiDou B1I signal [24], the code rate is 2.046MHz and the ranging of a single chip is about 147m. The length of the simulated data is 2000ms, the correlation time is 2ms, and the sampling frequency is 6MHz. A 12-element ULA with half-wavelength is used, and the tapped delay lines number is 5.

**TABLE 1.** Satellite signal paramter setting.

| PRN | Signal | Incident Angle (Degree) | $C/N_0$ (dBHz) | Delay (Chips) |
|---|---|---|---|---|
| 1 | Direct signal | 24 | 42 | 0 |
| 2 | Direct signal | 10 | 42 | 0 |
| | Multipath 1 | 17 | 39 | 0.1 |
| | Multipath 2 | -25 | 37 | 0.2 |
| | Multipath 3 | 72 | 35 | 0.2 |

**TABLE 2.** Interference paramter setting.

| Interference | Incident Angle (Degree) | Interference-to-noise ratio (dB) | Bandwidth (MHz) |
|---|---|---|---|
| 1 | 75 | 39 | 4 |
| 2 | -80 | 45 | 4 |
| 3 | -60 | 42 | 4 |

The parameters of the navigation signal and the interference are listed in Table 1 and Table 2, respectively. Two scenarios are simulated in this part, i.e. the scenarios with and without interference presence.

To verify the validity of the proposed method, it is compared with the space-time MVDR method, improved space-time MVDR method [6], and space-time PI method. Among them, the improved space-time MVDR method can guarantee a distortionless response for the direct signal. In this section, the space-time MVDR method, the improved space-time MVDR method, the space-time PI method, and the proposed method are called "MVDR", "IMVDR" "PI", and "Proposed", respectively.

### A. SIMULATION A

The scenario without interference is simulated in this part, and only the simulation results of the PRN2 is shown.

Fig. 5 shows the array pattern comparison results of different methods, in which the square stands for the direct signal component and the circles stand for the multipath components. As illustrated in Fig. 5, the PI method not only cannot enhance the satellite signal, but may cause energy cancellation. As for MVDR and IMVDR method, it can steer the array pattern gain to the direct signal, which can enhance the direct signal energy, but cannot null the multipath. Additionally, when the multipath is close to the direct signal, the multipath will also be enhanced. The proposed method can estimate the steering vector of the direct signal and multipath precisely without the help of the prior information of the direction of the direct signal, after beamforming method, the direct signal is enhanced and the direction of the multipath components is nulled.

The pseudo-range measurement is calculated by tracking the ACF of the signal in GNSS receiver. Ideally, the ACF is symmetry. However, the presence of multipath will cause the ACF distortion and unsymmetry, which will induce bias in pseudo-range measurement. Fig. 6 shows the normalized ACF results of different methods for PRN2. From Fig. 6, it can be found that the multipath has a much more influence on the distortion of the ACF for PI method than the MVDR method and IMVDR method. However, the ACF after the MVDR and IMVDR still have small distortions because the

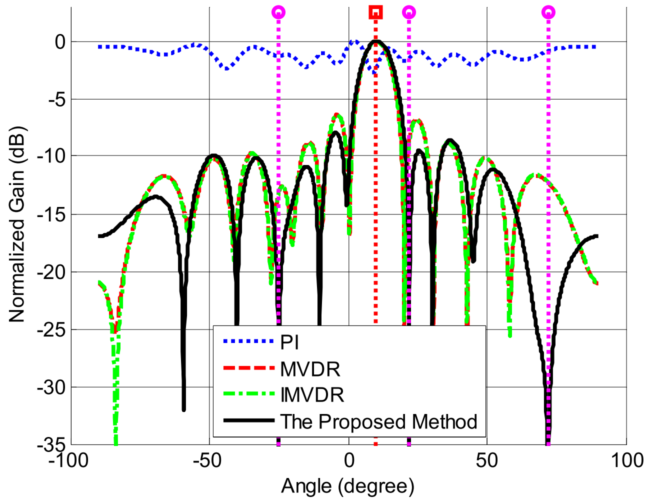


**FIGURE 5.** Normalized array gain patterns of different method for PRN2.

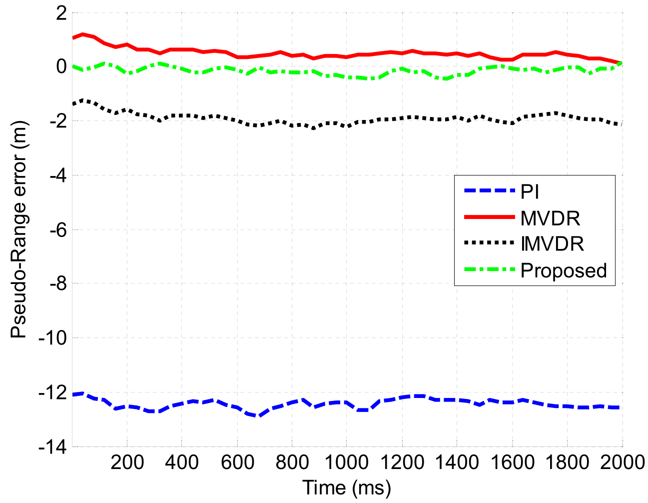


**FIGURE 7.** Pseudo-range error of different method for PRN2.

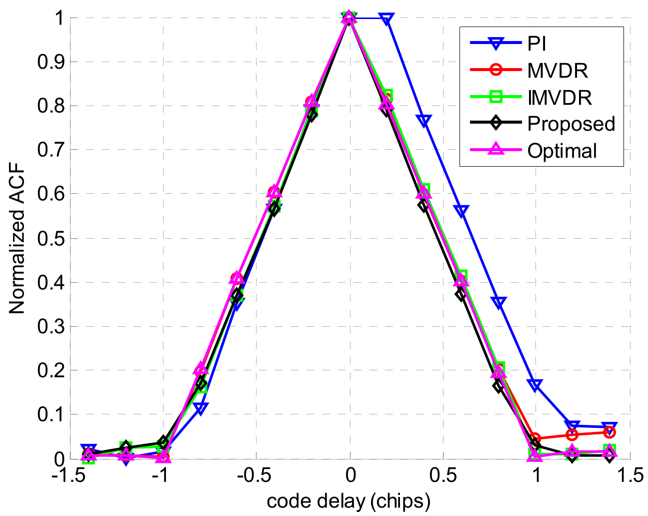


**FIGURE 6.** Normalized ACF results of different methods for PRN2.

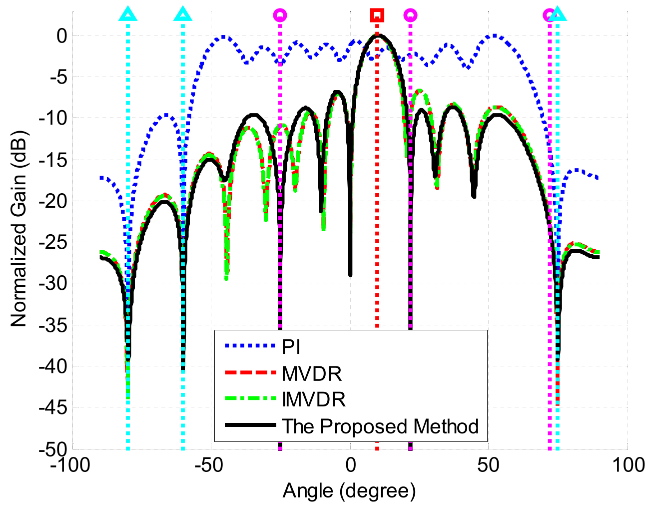


**FIGURE 8.** Normalized array gain patterns of different method.

multipath is not be suppressed completely. After multipath suppression using the proposed method, the ACF is well corrected, which can achieve the unbiased pseudo-range measurement.

In order to verify the influence of different methods, the pseudo-range measurement error is shown in Fig. 7. It can be seen from the figure that the pseudo-range error of PI method is larger than $-12$m. By enhancing the direct signal and increasing the multipath-to-signal ratio, the multipath error of the MVDR and IMVDR methods can be reduced to 1 or 2 meters. The proposed method can realize the unbiased pseudo-range measurement, which satisfies the needs of high precision applications.

## B. SIMULATION B

The scenario with interference is simulated in this part.

Fig. 8 shows the array pattern comparison results of PRN 2 for different methods, in which the triangles stand for the interference components. As illustrated in Fig. 8, all of the four methods can suppress the interference efficiently. The performance of multipath suppression using the traditional method is almost the same with the results that shown in simulation A. Additionally, if the incident angle of the multipath is close to the interference, the multipath will also be suppressed by the spatial-time filtering. Noting that the energy of the third multipath component is very low after interference suppression, it cannot be detected by the proposed method, which has little effect on the performance of multipath suppression.

Fig. 9 shows the normalized ACF results of different methods for PRN2. Both the space-time filtering and the multipath will cause the distortion to ACF, and the two parts cannot be separated. From Fig. 9, the distortion caused by the PI algorithm is greater than MVDR method and IMVDR method. The pseudo-range error results are shown in Fig. 10. The pseudo-range error after PI method is about $-8$m, and the error of MVDR and IMVDR is about 2m. The pseudo-range error is almost zero when using the proposed method.

In order to analyze the bias caused by space-time filtering, the normalized ACF and pseudo-range error of PRN1 are

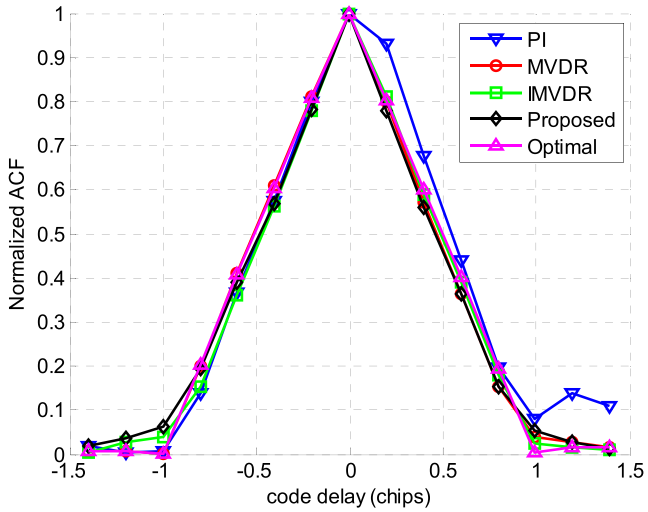


FIGURE 9. Normalized ACF results of different methods for PRN2.

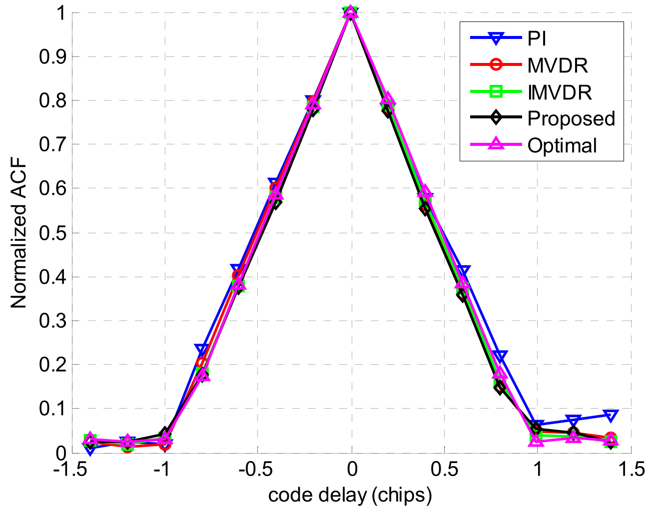


FIGURE 11. Normalized ACF results of different methods for PRN1.

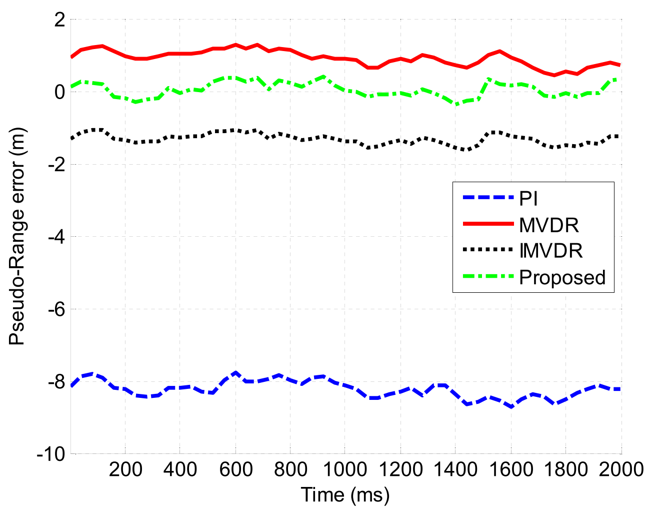


FIGURE 10. Pseudo-range error of different method for PRN2.

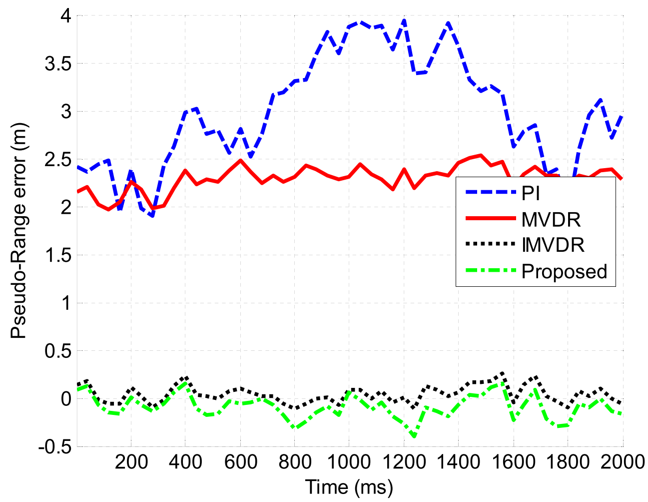


FIGURE 12. Pseudo-range error of different method for PRN1.

shown in Fig. 11 and 12, respectively. There are no multipath components for PRN1 signal, therefore, the error is mainly affected by the space-time filter. It can be seen from the figures that both the space-time PI and the space-time MVDR method will cause the ACF distortion and lead to bias in pseudo-range measurement. However, the proposed method and the IMVDR method can guarantee the distortionless response of the satellite signal, which leads to an unbiased pseudo-range measurement.

In summary, some conclusions can be drawn as follows.

(1) Space-time PI and MVDR method can cause the distortion of the satellite signal, leading to a bias in pseudo-range measurement, while the IMVDR method can achieve distortionless response of the satellite signal.

(2) When multipath is present, the PI method, MVDR method, and IMVDR method cannot steer the array pattern to null the direction of the multipath. MVDR and IMVDR method can reduce the influence of the multipath error by enhancing the direct signal energy.

(3) The proposed method can achieve interference and multipath suppression by using a cascaded beamforming method. The response of the satellite signal is distortionless and the pseudo-range measurement is unbiased. The multipath signal can be suppressed even with the absence of the interference. Furthermore, the direct signal is enhanced without using the prior knowledge of the direct signal direction.

## VI. CONCLUSION

A novel technique was proposed to suppress the interference and multipath by using a cascaded process method in GNSS receiver. In this method, the interference is firstly suppressed by subspace projection method. In the second stage, a multipath suppression algorithm based on sparse recovery using antenna array was proposed. The steer vectors of the direct signal and multipath signal are estimated using sparse recovery method. And the array antenna is used for beamforming to steer the gain in the direct signal direction and achieve a null in the multipath direction, so that the multipath suppression can be well realized. The bias caused by the space-time filter and the multipath error are mitigated efficiently without using the direction of the direct signal. Simulations demonstrate that the bias and multipath error can be mitigated successfully in the environment with or without interference, and unbiased pseudo-range measurement for precision application is provided.

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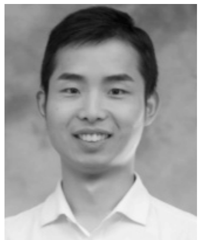

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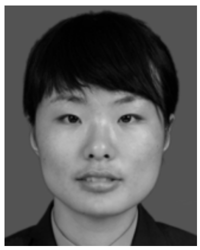

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