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Research Article

The PARAFAC-MUSIC Algorithm for DOA Estimation with Doppler Frequency in a MIMO Radar System

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The PARAFAC-MUSIC algorithm is proposed to estimate the direction-of-arrival (DOA) of the targets with Doppler frequency in a monostatic MIMO radar system in this paper. To estimate the Doppler frequency, the PARAFAC (parallel factor) algorithm is firstly utilized in the proposed algorithm, and after the compensation of Doppler frequency, MUSIC (multiple signal classification) algorithm is applied to estimate the DOA. By these two steps, the DOA of moving targets can be estimated successfully. Simulation results show that the proposed PARAFAC-MUSIC algorithm has a higher accuracy than the PARAFAC algorithm and the MUSIC algorithm in DOA estimation.

1. Introduction

Recently, there has been a growing interest in multiple-input multiple-output radar which utilizes multiple antennas at both transmitter and receiver. According to the array configuration, MIMO radar can be classified into two main types, the first of which is called distributed MIMO radar. It is composed of widely separated transmit antennas which transmit the linearly independent signals. Based on the sufficient distribution space and linearly independent signals, the distributed MIMO radar is able to obtain rich scattering properties of the targets and to mitigate radar cross-section (RCS) fluctuations. The second type is called collocated MIMO radar, in which the transmit elements are collocated. By transmitting independent waveforms and capitalizing on the MIMO spatial signature, the collocated MIMO can estimate the parameters of interest via coherent processing.

On the other hand, an accurate estimation of signal DOA has made a great sense in radar system of military and commercial application. Many algorithms were proposed on DOA estimation in MIMO radar system as it provides waveform diversity and spatial distribution of flexibility. In [1], the Capon algorithm, proposed in 1969, is a main lobe self-adaptive algorithm which is stable but poor at low SNR. ESPRIT (estimation signal parameter via rotational invariance techniques) algorithm in [2] exploits the invariance property for DOA estimation in MIMO radar system. In [3], MUSIC algorithm, one of the most popular algorithms for DOA estimation, utilizes subspace analysis and has a good performance.

Although these algorithms introduced above are useful for DOA estimation, they are poor when the Doppler frequency is taken into consideration. As the target always moves, DOA estimation of moving targets is extremely essential. In [4, 5], PARAFAC algorithm exploits the iteration of TALS (trilinear alternating least square) to estimate the frequency and DOA. PARAFAC algorithm has a better performance than Capon, ESPRIT, and MUSIC when considering Doppler frequency. PARAFAC-MUSIC algorithm, proposed in this paper, utilizes the PARAFAC algorithm to estimate the Doppler frequency and then exploits MUSIC algorithm to estimate the DOA after the Doppler effect is eliminated. Simulation shows that PARAFAC-MUSIC algorithm is also able to solve the problem generated from Doppler frequency and may have a better performance than PARAFAC algorithm.

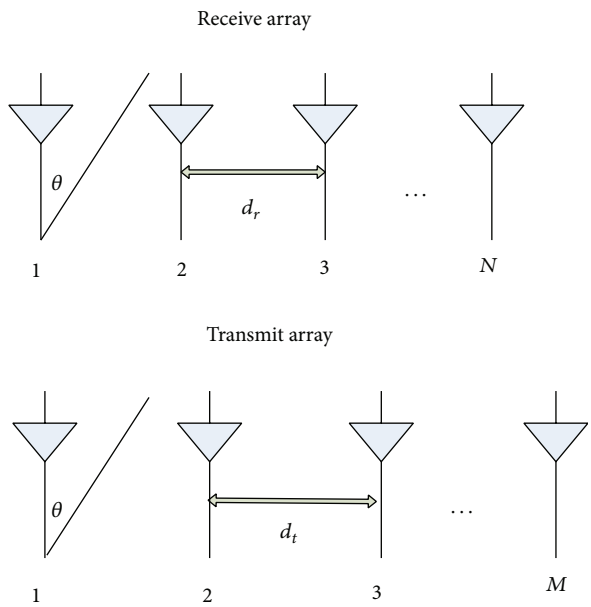

Figure 1: Transmit and receive antennas.

2. Signal Model

Assume that there is a monostatic MIMO radar system with M collocated transmit antennas and N collocated receive antennas, both of which are uniform linear arrays (ULA), shown in Figure 1. Since the transmit array and the receive array are assumed to be close to each other, the DOD (direction of departure) and the DOA are approximately equal, denoted by θ. d_t and d_r are the interelement spacing at the transmitter and receiver. The M transmit antennas transmit orthogonal waveforms simultaneously while the N receive antennas receive the signals reflected from the targets. It is supposed that there are P targets in the far field, each of which can be considered as a point source. And the RCS of each target is assumed as a constant during a pulse period.

The received signal at time t before match filters can be expressed as

$$\mathbf{x}(t) = \sum_{p=1}^{P} \mathbf{a}_r(\theta_p) \times \mathbf{a}_t(\theta_p)^T \times \mathbf{s}_p(t) + \mathbf{n}_{\mathbf{x}}(t), \tag{1}$$

where $\mathbf{a}_r(\theta_p) = [1, e^{-j2\pi d_r \sin\theta_p/\lambda}, \ldots, e^{-j2\pi(N-1)d_r \sin\theta_p/\lambda}]^T$ is the receive steering vector and $\mathbf{a}_t(\theta_p) = [1, e^{-j2\pi d_t \sin\theta_p/\lambda}, \ldots, e^{-j2\pi(M-1)d_t \sin\theta_p/\lambda}]^T$ is the transmit steering vector. λ denotes the wavelength. Consider $\mathbf{s}_p(t) = [\beta_p e^{-j2\pi f_{dp} t} s_1(t), \ldots, \beta_p e^{-j2\pi f_{dp} t} s_M(t)]^T$, where β_p is the reflection coefficient depending on the RCS of the target p, f_{dp} is Doppler frequency of the target p, and $s_m(t)$ is the transmit signal of each transmit antenna. $\mathbf{n}_{\mathbf{x}}(t)$ is a complex Gaussian white noise vector with zero mean and covariance matrix $\sigma^2\mathbf{I}$. $[\cdot]^T$ represents transpose operator.

As the transmit waveforms are orthogonal from each other, the output of the match filters at time t is

$$\mathbf{y}(t) = [\mathbf{A}_t(\theta) \odot \mathbf{A}_r(\theta)]\, c(t) + \mathbf{n}_{\mathbf{y}}(t), \tag{2}$$

where

$$\begin{aligned} \mathbf{A}_t(\theta) &= [\mathbf{a}_t(\theta_1), \ldots, \mathbf{a}_t(\theta_P)], \\ \mathbf{A}_r(\theta) &= [\mathbf{a}_r(\theta_1), \ldots, \mathbf{a}_r(\theta_P)], \end{aligned} \tag{3}$$

$\mathbf{c}(t) = [\beta_1 e^{j2\pi f_{d1} t}, \ldots, \beta_P e^{j2\pi f_{dP} t}]^T$, and "$\odot$" represents the Khatri-Rao product.

So the received signal before the match filters can be expressed by matrix as

$$\mathbf{X} = \sum_{p=1}^{P} \left[\mathbf{a}_r(\theta_p) \times \mathbf{a}_t(\theta_p)^T \times \mathbf{S}\right] + \mathbf{W}_{\mathbf{X}}, \tag{4}$$

where $\mathbf{S} = [\mathbf{s}_1, \mathbf{s}_2, \mathbf{s}_3, \ldots, \mathbf{s}_L]$ and L denotes the number of snapshot. $\mathbf{W}_{\mathbf{X}}$ is complex Gaussian white noise matrix.

The output at match filters can be expressed by matrix as

$$\mathbf{Y} = [\mathbf{A}_t(\theta) \odot \mathbf{A}_r(\theta)]\, \mathbf{C}^T + \mathbf{W}_{\mathbf{Y}}, \tag{5}$$

where $\mathbf{Y} = [\mathbf{y}_1, \mathbf{y}_2, \mathbf{y}_3, \ldots, \mathbf{y}_L]$ is a $MN \times L$ matrix and $\mathbf{C}^T = [c_1, c_2, c_3, \ldots, c_L]$ is a $P \times L$ matrix. Consider $c_l = [\beta_1 e^{-j2\pi f_{d1} lT}, \ldots, \beta_P e^{-j2\pi f_{dP} lT}]^T$, in which T is sampling time. $\mathbf{W}_{\mathbf{Y}}$ is complex Gaussian white noise matrix.

3. PARAFAC-MUSIC Algorithm

PARAFAC-MUSIC algorithm combines the advantages of PARAFAC algorithm and MUSIC algorithm. It exploits PARAFAC algorithm to estimate the Doppler frequency, then eliminates the effect in received signal generated from Doppler frequency, and finally utilizes MUSIC algorithm to estimate the DOA.

As known from formula (5), the output $\mathbf{Y}$ at match filters possesses the character of three-way model. Therefore, it can be expressed as set $\mathbf{Y}_L$ composed of L sections $\underset{\cdot}{\mathbf{Y}}_l$, each of which is a $M \times N$ matrix. $\underset{\cdot}{\mathbf{Y}}_l$ can be expressed as

$$\underset{\cdot}{\mathbf{Y}}_l = \mathbf{A}_t(\theta) \times D_l[\mathbf{C}] \times \mathbf{A}_r(\theta)^T + \underset{\cdot}{\mathbf{w}}_l, \tag{6}$$

where $D_j[\cdot]$ represents the diagonal matrix composed of the elements obtained from the jth row of the matrix.

In the same way, $\mathbf{Y}$ can be expressed as a set $\mathbf{Y}_N$ composed of N sections $\underset{\cdot\cdot}{\mathbf{Y}}_n$ or a set $\mathbf{Y}_M$ composed of M sections $\underset{\cdot\cdot}{\mathbf{Y}}_m$. $\underset{\cdot}{\mathbf{Y}}_n$ and $\underset{\cdot\cdot}{\mathbf{Y}}_m$ can be, respectively, expressed as

$$\begin{aligned} \underset{\cdot\cdot}{\mathbf{Y}}_n &= \mathbf{C} \times D_n[\mathbf{A}_r(\theta)] \times \mathbf{A}_t(\theta)^T + \underset{\cdot\cdot}{\mathbf{w}}_n, \\ \underset{\cdot\cdot}{\mathbf{Y}}_m &= \mathbf{A}_r(\theta) \times D_m[\mathbf{A}_t(\theta)] \times \mathbf{C}^T + \underset{\cdot\cdot}{\mathbf{w}}_m. \end{aligned} \tag{7}$$

So $\mathbf{Y}_L$, $\mathbf{Y}_N$, and $\mathbf{Y}_M$ can be, respectively, expressed as

$$\begin{aligned}
\mathbf{Y}_L &= \begin{bmatrix} \underset{\cdot}{\mathbf{Y}}_1 \\ \underset{\cdot}{\mathbf{Y}}_2 \\ \vdots \\ \underset{\cdot}{\mathbf{Y}}_L \end{bmatrix} \\
&= \begin{bmatrix} \mathbf{A}_t(\theta) \times D_1[\mathbf{C}] \\ \mathbf{A}_t(\theta) \times D_2[\mathbf{C}] \\ \vdots \\ \mathbf{A}_t(\theta) \times D_L[\mathbf{C}] \end{bmatrix} \mathbf{A}_r(\theta)^T + \begin{bmatrix} \underset{\cdot}{\mathbf{w}}_1 \\ \underset{\cdot}{\mathbf{w}}_2 \\ \vdots \\ \underset{\cdot}{\mathbf{w}}_L \end{bmatrix} \\
&= [\mathbf{C} \odot \mathbf{A}_t(\theta)]\, \mathbf{A}_r(\theta)^T + \mathbf{W}_L, \\
\mathbf{Y}_N &= \begin{bmatrix} \underset{\cdot\cdot}{\mathbf{Y}}_1 \\ \underset{\cdot\cdot}{\mathbf{Y}}_2 \\ \vdots \\ \underset{\cdot\cdot}{\mathbf{Y}}_N \end{bmatrix} \\
&= \begin{bmatrix} \mathbf{C} \times D_1[\mathbf{A}_r(\theta)] \\ \mathbf{C} \times D_2[\mathbf{A}_r(\theta)] \\ \vdots \\ \mathbf{C} \times D_N[\mathbf{A}_r(\theta)] \end{bmatrix} \mathbf{A}_t(\theta)^T + \begin{bmatrix} \underset{\cdot\cdot}{\mathbf{w}}_1 \\ \underset{\cdot\cdot}{\mathbf{w}}_2 \\ \vdots \\ \underset{\cdot\cdot}{\mathbf{w}}_N \end{bmatrix} \\
&= [\mathbf{A}_r(\theta) \odot \mathbf{C}]\, \mathbf{A}_t(\theta)^T + \mathbf{W}_N, \\
\mathbf{Y}_M &= \begin{bmatrix} \underset{\cdot\cdot\cdot}{\mathbf{Y}}_1 \\ \underset{\cdot\cdot\cdot}{\mathbf{Y}}_2 \\ \vdots \\ \underset{\cdot\cdot\cdot}{\mathbf{Y}}_M \end{bmatrix} \\
&= \begin{bmatrix} \mathbf{A}_r(\theta) \times D_1[\mathbf{A}_t(\theta)] \\ \mathbf{A}_r(\theta) \times D_2[\mathbf{A}_t(\theta)] \\ \vdots \\ \mathbf{A}_r(\theta) \times D_M[\mathbf{A}_t(\theta)] \end{bmatrix} \mathbf{C}^T + \begin{bmatrix} \underset{\cdot\cdot\cdot}{\mathbf{w}}_1 \\ \underset{\cdot\cdot\cdot}{\mathbf{w}}_2 \\ \vdots \\ \underset{\cdot\cdot\cdot}{\mathbf{w}}_M \end{bmatrix} \\
&= [\mathbf{A}_t(\theta) \odot \mathbf{A}_r(\theta)]\, \mathbf{C}^T + \mathbf{W}_M.
\end{aligned} \tag{8}$$

When the three-way array is obtained, it is common to utilize TALS [5] to estimate $\widehat{\mathbf{A}}_t(\theta)$, $\widehat{\mathbf{A}}_r(\theta)$, and $\widehat{\mathbf{C}}$. TALS is a popular method in data detection of three-way array model which updates one estimated matrix in each step after obtaining the initial estimated matrixes. Least square (LS) algorithm always works in estimated matrix update that takes the estimated matrix to be updated as a variable and other estimated matrixes as constants in each step. When all the estimated matrixes are updated, it will carry on the next iteration until convergence. Specific steps of TALS are stated as follows.

Step 1. Construct the three-way array $\mathbf{Y}_L$, $\mathbf{Y}_N$, $\mathbf{Y}_M$ based on the output at match filters.

Step 2. Initialize all the estimated matrixes $\widehat{\mathbf{A}}_{t0}(\theta)$, $\widehat{\mathbf{A}}_{r0}(\theta)$, and $\widehat{\mathbf{C}}_0$.

Step 3. Substitute $\widehat{\mathbf{A}}_{t(k-1)}(\theta)$, $\widehat{\mathbf{C}}_{k-1}$ into formula (9) to obtain $\widehat{\mathbf{A}}_{rk}(\theta)$ shown as formula (10). $[\cdot]_F$, # represents Frobenius norm and the pseudo inverse operation. k represents the iterations

$$\widehat{\mathbf{A}}_{rk}(\theta) = \arg\min_{\widehat{\mathbf{A}}_r} \left\| \mathbf{Y}_L - \left[\widehat{\mathbf{C}}_{k-1} \odot \widehat{\mathbf{A}}_{t(k-1)}(\theta)\right] \widehat{\mathbf{A}}_{rk}^T(\theta) \right\|_F^2, \tag{9}$$

$$\widehat{\mathbf{A}}_{rk}(\theta) = \mathbf{Y}_L^T \left[\left(\widehat{\mathbf{C}}_{k-1} \odot \widehat{\mathbf{A}}_{t(k-1)}(\theta) \right)^{\#} \right]^T. \tag{10}$$

Step 4. Substitute $\widehat{\mathbf{A}}_{rk}(\theta)$, $\widehat{\mathbf{C}}_{k-1}$ into formula (11) to obtain $\widehat{\mathbf{A}}_{tk}(\theta)$ shown as formula (12). Consider

$$\widehat{\mathbf{A}}_{tk}(\theta) = \arg\min_{\widehat{\mathbf{A}}_t} \left\| \mathbf{Y}_N - \left[\widehat{\mathbf{A}}_{rk}(\theta) \odot \widehat{\mathbf{C}}_{k-1}\right] \widehat{\mathbf{A}}_{tk}^T(\theta) \right\|_F^2, \tag{11}$$

$$\widehat{\mathbf{A}}_{tk}(\theta) = \mathbf{Y}_N^T \left[\left(\widehat{\mathbf{A}}_{rk}(\theta) \odot \widehat{\mathbf{C}}_{k-1}^T \right)^{\#} \right]^T. \tag{12}$$

Step 5. Substitute $\widehat{\mathbf{A}}_{tk}(\theta)$, $\widehat{\mathbf{A}}_{rk}(\theta)$ into formula (13) to obtain $\widehat{\mathbf{C}}_k$ shown as formula (14). Then calculate $\delta_k = \sum_{l=1}^{L} \|\mathbf{Y}_l - \widehat{\mathbf{A}}_{tk}(\theta) D_l[\mathbf{C}] \widehat{\mathbf{A}}_{rk}(\theta)^T\|_F^2$. If $|\delta_k - \delta_{k-1}| > \varepsilon$ (ε is error threshold), repeat Step 3 to Step 5; otherwise, go to Step 6. Consider

$$\widehat{\mathbf{C}}_k = \arg\min_{\widehat{\mathbf{C}}} \left\| \mathbf{Y}_M - \left[\widehat{\mathbf{A}}_{tk}(\theta) \odot \widehat{\mathbf{A}}_{rk}(\theta)\right] \widehat{\mathbf{C}}_k^T \right\|_F^2, \tag{13}$$

$$\widehat{\mathbf{C}}_k = \mathbf{Y}_M^T \left[\left(\widehat{\mathbf{A}}_{tk}(\theta) \odot \widehat{\mathbf{A}}_{rk}(\theta) \right)^{\#} \right]^T. \tag{14}$$

Step 6. $\widehat{\mathbf{A}}_t(\theta)$, $\widehat{\mathbf{A}}_r(\theta)$, and $\widehat{\mathbf{C}}$ are obtained after iteration. Then $\widehat{f}_{dp}$ can be calculated by the following formula:

$$\widehat{f}_{dp} = \frac{1}{2\pi T(L-1)} \sum_{i=1}^{L-1} \operatorname{angle}\left(\frac{\widehat{c}_{i,p}(\theta)}{\widehat{c}_{i+1,p}(\theta)} \right), \tag{15}$$

where $\widehat{c}_{i,p}(\theta)$ presents the element in $\widehat{\mathbf{C}}^T$ and i and p are the indexes of row and column. angle$(\cdot)$ represents phase obtain operation.

When the Doppler frequency has been estimated, it can be used to eliminate the Doppler effect to received signal $\mathbf{X}$. The signal after preprocessing can be expressed as

$$\overleftrightarrow{\mathbf{X}} = \left[\overleftrightarrow{\mathbf{x}}_1, \overleftrightarrow{\mathbf{x}}_2, \overleftrightarrow{\mathbf{x}}_3, \ldots, \overleftrightarrow{\mathbf{x}}_L \right], \tag{16}$$

where $\overleftrightarrow{\mathbf{x}}_l = \overleftrightarrow{\mathbf{x}}(lT)$, $\overleftrightarrow{\mathbf{x}}(t) = \sum_{p=1}^{P} [\mathbf{x}_p(t)/f_p(t)]$. $\mathbf{x}_p(t) = \mathbf{a}_r(\theta_p) \times \mathbf{a}_t(\theta_p)^T \times \mathbf{s}_p(t) + \mathbf{n}_p(t)$, $f_p(t)$ represents the compensatory phase of target p and can be expressed as

$$f_p(t) = e^{-j2\pi \widehat{f}_{dp} t}. \tag{17}$$

So the output of match filters can be expressed as

$$\overline{\mathbf{X}} = \operatorname{vec}\left(\frac{\overleftrightarrow{\mathbf{X}} \mathbf{S}^H}{L} \right), \tag{18}$$

where $[\cdot]^H$ represents Hermit operation and vec$(\cdot)$ represents vector obtain operation.

The covariance of $\overline{\mathbf{X}}$ can be expressed as

$$\mathbf{R}_{\overline{\mathbf{X}}\overline{\mathbf{X}}} = E\left[\overline{\mathbf{X}}\,\overline{\mathbf{X}}^H\right]. \tag{19}$$

Then eigendecomposition is carried on for $\mathbf{R}_{\overline{\mathbf{X}}\overline{\mathbf{X}}}$ shown in formula (20). Consider

$$\mathbf{R}_{\overline{\mathbf{X}}\overline{\mathbf{X}}} = \mathbf{U}\sum\mathbf{U}^H, \tag{20}$$

where $\sum = \mathrm{diag}(\alpha_1, \alpha_2, \ldots, \alpha_{M\times N})$ and α_{ij} denotes eigenvalue of $\mathbf{R}_{\overline{\mathbf{X}}\overline{\mathbf{X}}}$.

Since the first p eigenvalues are composed of variance of signal and Gaussian white noise while the remaining eigenvalues are only composed of variance of Gaussian white noise, the first p eigenvalues are larger than the rest at high SNR. The first p eigenvalues are defined as "signal eigenvalues" and the rest are defined as "noise eigenvalues." So the eigenvector matrix $\mathbf{U}$ can be classified into two parts in the following formula:

$$\mathbf{U} = \left[\mathbf{G}_s \mid \mathbf{G}_n\right], \tag{21}$$

where $\mathbf{G}_s = [\mathbf{u}_1, \mathbf{u}_2, \ldots, \mathbf{u}_p]$ is composed of the signal eigenvectors and $\mathbf{G}_n = [\mathbf{u}_{p+1}, \mathbf{u}_{p+2}, \ldots, \mathbf{u}_{M\times N}]$ is composed of the noise eigenvectors.

So the spectrum estimation formula of MUSIC is expressed as

$$P_{\mathrm{MUSIC}}(\theta) = \frac{1}{\boldsymbol{\omega}(\theta)^H\mathbf{G}_n\mathbf{G}_n^H\boldsymbol{\omega}(\theta)}, \tag{22}$$

where $\boldsymbol{\omega}(\theta) = \mathbf{a}_r(\theta) \otimes \mathbf{a}_t(\theta)$. "$\otimes$" represents the Kronecker product.

Then the estimated DOA can be obtained according to the position of the spectral peak.

4. Results and Discussion

In order to state that PARAFAC-MUSIC algorithm has a better performance in DOA estimation with Doppler frequency, simulation is carried on with MATLAB software to compare PARAFAC-MUSIC algorithm with PARAFAC algorithm and MUSIC algorithm.

Assume that there are two targets in the far field and a MIMO radar system with 8 collocated transmit antennas and 8 collocated receive antennas. The frequency of the transmit signals is 4 GHz. The array structure is the same as Figure 1 with $d_t = 0.5\lambda\times M$ space between adjacent transmit elements and $d_r = 0.5\lambda$ space between adjacent receive elements. The targets are at 5° and 15°, respectively, relative to the MIMO radar system with the same velocity $v_t = 300$ m/s and the same RCS $\beta = 1\ \mathrm{m}^2$. The sampling frequency is 100 KHz. The number of snapshots is $L = 80$. Additionally, results shown below are obtained from 500 Monte Carlo experiments at each SNR.

Figures 2 and 3 show the two targets' mean errors of the estimated DOAs obtained from PARAFAC algorithm, MUSIC algorithm, and proposed algorithm. It is apparent to find that the mean error of target 1 obtained from MUSIC algorithm fluctuates near 0.36° and converges on 0.36° gradually with the increasing SNR, and the mean error of target 2 obtained from MUSIC algorithm fluctuates near 0.31° and converges on 0.31° gradually with the increasing SNR. The results show that the MUSIC algorithm cannot provide an accurate estimation because of the Doppler frequency. Nevertheless, as far as the PARAFAC algorithm and the proposed algorithm are concerned, the mean error of two targets fluctuates near 0° and converges on 0° gradually with

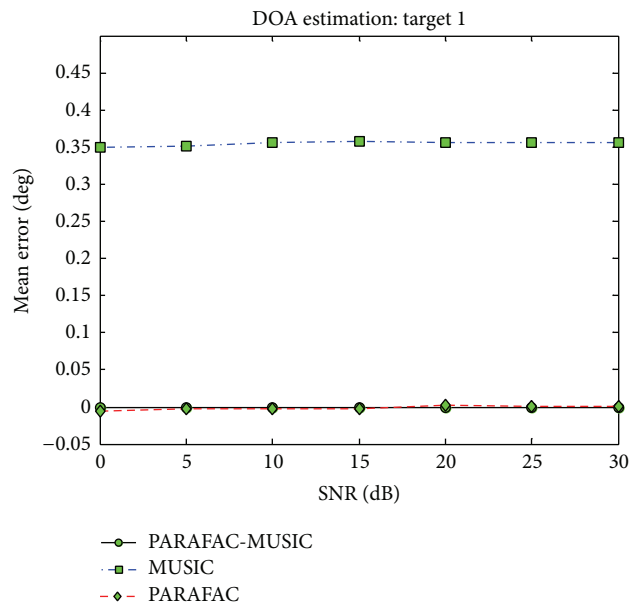

Figure 2: Mean error of target 1.

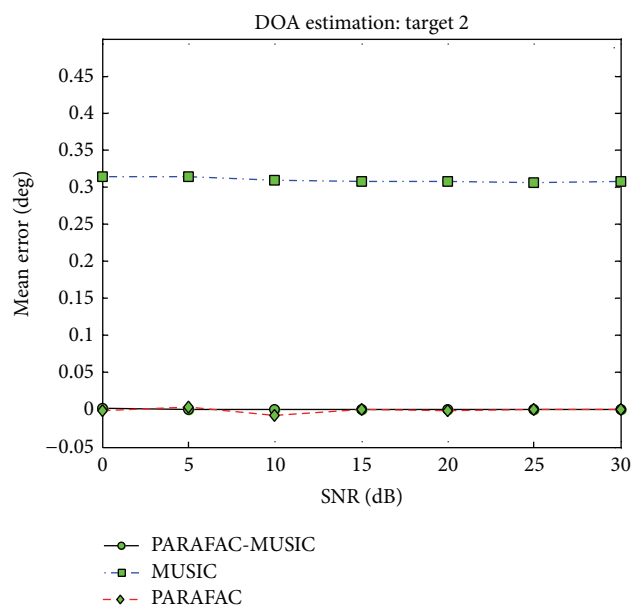

Figure 3: Mean error of target 2.

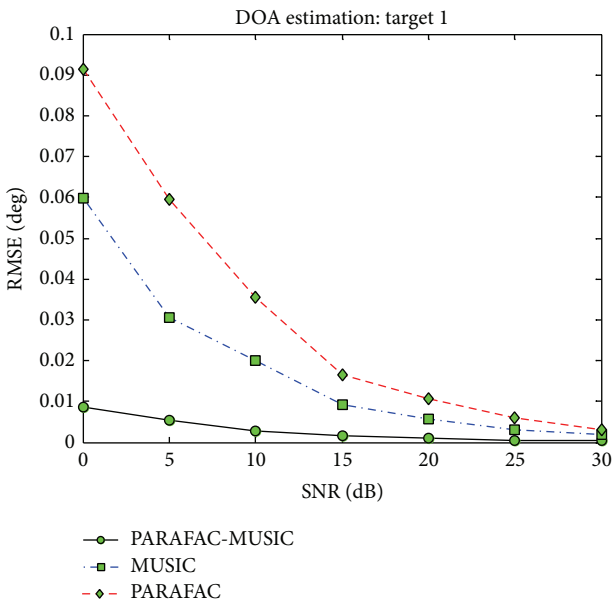

Figure 4: RMSE of target 1.

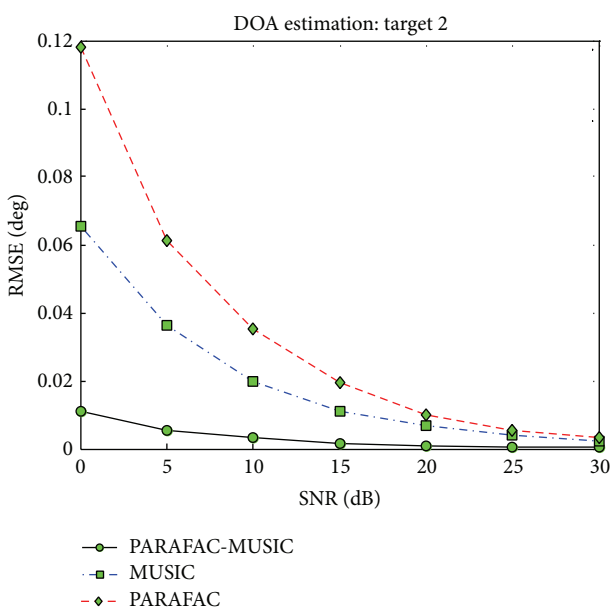

Figure 5: RMSE of target 2.

the increasing SNR. But the mean error of the proposed algorithm in both figures is closer to 0° than the one of PARAFAC algorithm which means that the proposed algorithm has a better performance than the PARAFAC algorithm.

Figures 4 and 5 illustrate the two targets' root mean square errors (RMSE) of the estimated DOAs obtained from PARAFAC algorithm, MUSIC algorithm, and proposed algorithm. RMSE of the estimated DOAs in both figures decreases gradually with the increasing SNR. Though the RMSE of the estimated DOAs in both figures obtained from the MUSIC algorithm, the PARAFAC algorithm and the proposed algorithm tend to 0° with the increasing SNR; the RMSE of the proposed algorithm is smaller than the other two algorithms which means that the proposed algorithm has a smaller fluctuation in DOA estimation.

Simulation results state that MUSIC algorithm suffers from Doppler frequency seriously and cannot provide an accurate DOA estimation to moving target. Besides, the PARAFAC-MUSIC is able to estimate the DOA with a higher accuracy than the PARAFAC algorithm to moving targets.

5. Conclusions

In this paper, some algorithms in DOA estimation are reviewed first. Then a new algorithm named PARAFAC-MUSIC is proposed to estimate the DOA with Doppler frequency in a MIMO radar system. The proposed algorithm is able to estimate DOAs of multiple targets. Finally simulation results confirm that the MUSIC algorithm is not suitable for DOA estimation with Doppler frequency while PARAFAC-MUSIC algorithm is very popular for DOA estimation with Doppler frequency and has a higher accuracy than the PARAFAC algorithm and the MUSIC algorithm.

Conflict of Interests

The authors declare that there is no conflict of interests regarding the publication of this paper.

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