Two-Dimensional Angle Estimation for Monostatic MIMO Radar Using Expanded PARAFAC Model

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Abstract — In this research, we address the problem of Two-Dimensional (2D) angle estimation for monostatic Multiple-Input Multiple-Output (MIMO) radar. An expanded PARAFAC model is proposed to make full use of the Vandermonde-like structure of the data model, which expands the received data through unitary transformation and links the problem of 2D angle estimation to the PARAFAC model. Unlike the traditional estimation algorithms such as multiple signal classification (MUSIC) and estimation method of signal parameters via rotational invariance techniques (ESPRIT), the proposed algorithm does not require spectral peak searching nor eigenvalue decomposition of the received signal covariance matrix. Furthermore, our algorithm can achieve automatic pairing of 2D angles, and it has blind and robust characteristic, therefore the proposed algorithm has higher working efficiency. In addition, the proposed algorithm can detect more targets and has better estimation accuracy than ESPRIT algorithm and PARAFAC method. Extensive numerical experiments verify the effectiveness and improvement of our algorithm.

Index Terms—multiple-input multiple-output radar, angle estimation, PARAFAC decomposition

I. INTRODUCTION

There is no doubt that Multiple-Input Multiple-Output (MIMO) radar will be a backbone to the future war. The concept of MIMO radar is that the radar system simultaneously transmits mutually orthogonal waveforms with multiple antennas and receive the reflected echoes with multiple antennas. The information of an individual transmitter-to-receiver path is separated by the matched filter at the receiving end. The virtual transmitter-to-receiver paths enable MIMO radar to achieve more degrees of freedom than the traditional phase-array radar [1]. Theoretical research indicates that MIMO radar has several built-in advantages in noise suppression, overcoming the fading effect, improving spatial resolution, enhancing parameter identifiability, etc. [2]-[4]. Direction-of-Arrival (DOA) estimation is a canonical problem in MIMO radar that has aroused extensive attention in the past decade. In [5] and [6], Capon and multiple signal classification (MUSIC) algorithms were introduced into angle estimation for MIMO radar, respectively, which estimate parameters via the spectral peak searching and suffer from high computational complexity. ESPRIT method, which is short for the estimation method of signal parameters via rotational invariance techniques, has been discussed in [7] and [8]. It utilizes the shift invariance property of the virtual array in MIMO radar for parameters estimation, and ESPRIT method does not require spectral peak searching. Generally speaking, the computational complexity of multiplying two complex numbers is much more than that of two real numbers. To reduce the computation load, the unitary transform based ESPRIT algorithm was proposed in [9] for target location in bistatic MIMO radar, which computing ESPRIT with real number and has estimation performance very close to that of the ESPRIT method. The parallel factor analysis (PARAFAC) method, sometimes also referred to as trilinear decomposition algorithm, was developed for angle estimation in [10]-[11], which stacked up the received data into higher-dimensional tensors model. Along with Trilinear Alternating Least Square (TALS) [12], the loading matrices could be estimated thus parameters were obtained. To support on-line applications, an adaptive PARAFAC algorithm has been presented in [13], which lower the complexity of the existing batch mode PARAFAC algorithms and has estimation performance very close to that of non-adaptive ones. The compressed PARAFAC model has been derived in [14], which fits a low-rank tensor model in the compressed domain and solves the inverse problem in the low-dimensional space. It has been applied in angle and Doppler frequency estimation for monostatic uniform linear arrays (ULA) configured MIMO radar [15]. The PARAFAC method may be considered as a generalization of the ESPRIT method while achieving more accurate performance [16].

The ULA configured MIMO radar only have the ability for one-dimensional (1D) angle estimation, for Two-Dimensional (2D) angle estimation, 2D array manifold is needed. Typical 2D arrays including octagon array [17], cross array [18], L-shape array [19] and rectangular array [20]. It has been proven that an L-shape array has better accuracy potential than the other

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manifolds [21]. Inspired by the expansion effect of the unitary transform [11], an improved PARAFAC algorithm is developed for 2D angle estimation in this study. The virtual received data is formed through unitary transform, 2D angle estimation is linked to the PARAFAC model. The expanded PARAFAC model increase the effective aperture of the radar system, therefore the proposed algorithm outperformed the conventional PARAFAC algorithm [18]. Furthermore, it does not require singular value decomposition of the received data and can obtain automatically paired 2D angle estimations in the MIMO radar. Finally, computation complexity is analyzed and Cramér-Rao Bound (CRB) for 2D angle estimation in MIMO radar is derived. Also, simulation results are given to illustrate the good estimation capability of the proposed algorithm for L-shape monostatic MIMO radar system.

The paper outline is as follows. The data model for the L-shape monostatic MIMO radar is presented in Section 2. The expanded PARAFAC algorithm is derived in Section 3. The performance and complexity of the proposed algorithm is discussed in Section 4. Simulation results are given in Section 5. We end the paper by a brief concluding in Section 6.

Notation, capital letters $X$ and lower case $x$ in bold denote, respectively, matrices and vectors. The superscript $\transp$, $\herm$, $\herm^\transp$ and $\herm^\transp$ represent the operations of transpose, Hermitian transpose, inverse and pseudo-inverse, respectively; The subscript $\|\|_F$ denote the Frobenius norm of $X$; $\otimes$ stands for the Kronecker product; The Khatri-Rao product (column-wise matrix) is denoted by $\otimes$ , i.e., $[a_1 \otimes b_1, a_2 \otimes b_2, \ldots, a_k \otimes b_k]$; The $M \times M$ identity matrix is denoted by $I_M$, and the $M \times M$ inverse permutation matrix is denoted by $\Pi_M$.

II. DATA MODEL

We modeled a colocated MIMO radar system with L-shape configuration, as shown in Fig. 1. Suppose that there is an $M$-element transmitter located in the $x$-axis and an $N$-element receiver located in the $y$-axis, both of them are ULA with half-wavelength spacing, and the phase reference point of the transmitter and receiver are overlapped in the origin. Assume that $K$ non-coherent targets appearing in the far-field of the antennas with the $k$-th target at azimuth angle $\varphi_k$ and elevation angle $\theta_k$. Additional assumptions are that the transmit antennas transmit narrowband orthogonal waveforms with the same carrier frequency. The echoes are collected at each receive antenna and passed through a set of matched filters for information extraction. The output of the matched filters in the $n$-th receiver takes the form [22]

$$ r_n(t) = \sum_{k=1}^{K} a_k^*(\theta_k, \varphi_k) a_n(\theta_k, \varphi_k) s_k(t) $$

where

$$ a_n(\theta_k, \varphi_k) = \left[ a_n^*(\theta_k, \varphi_k), \ldots, a_n^*(\theta_k, \varphi_k) \right] \in \mathbb{C}^{N \times 1} $$

$$ a_k(\theta_k, \varphi_k) = \left[ a_k^*(\theta_k, \varphi_k), \ldots, a_k^*(\theta_k, \varphi_k) \right] \in \mathbb{C}^{M \times 1} $$

respectively, denote the receive steering vector and the transmit steering vector for the $k$-th target, with the element $a_n^*(\theta_k, \varphi_k) = \exp\left(-j\pi(n-1)\sin\theta_k \cos\varphi_k \right)$ $(n = 1, 2, \ldots, N)$, and the element $a_k^*(\theta_k, \varphi_k) = \exp\left(-j\pi(m-1)\sin\theta_k \sin\varphi_k \right)$ $(m = 1, 2, \ldots, M)$.

$s_k(t) = a_k \exp\left(j2\pi f_i t/f_s \right)$ is the product of the radar cross section (RCS) $a_k$ and the Doppler frequency shift $f_i/f_s$ of the $k$-th target. $f_i$ is the Doppler frequency, $f_s$ is the pulse repeat frequency. Arrange $x(t) = [r_1(t), r_2(t), \ldots, r_n(t)]^\transp$, we have

$$ x(t) = [a_1(\theta_1, \varphi_1) \otimes a_1(\theta_1, \varphi_1), \ldots, a_K(\theta_K, \varphi_K) \otimes a_K(\theta_K, \varphi_K)] s(t) $$

where $s(t) = [s_1(t), \ldots, s_K(t)]^\transp$ is a column vector. Let the transmit direction matrix $A_T = [a_1(\theta_1, \varphi_1) \ldots a_K(\theta_K, \varphi_K)] \in \mathbb{C}^{M \times K}$, the receive direction matrix $A_R = [a_1(\theta_1, \varphi_1) \ldots a_K(\theta_K, \varphi_K)] \in \mathbb{C}^{N \times K}$.

Define the coefficient matrix as $S = \begin{bmatrix} s(1), \ldots, s(L) \end{bmatrix}$\transp \in \mathbb{C}^{L \times K}$. Suppose that the directions of targets and the RCS coefficients are consist during $L$ snapshots, hence the coefficient matrix has a Vandermonde-like structure without considering the RCSs:

$$ S = \begin{bmatrix} a_1 \exp\left(j2\pi f_i/f_s \right) & \ldots & a_K \exp\left(j2\pi f_i/f_s \right) \\ a_1 \exp\left(j4\pi f_i/f_s \right) & \ldots & a_K \exp\left(j4\pi f_i/f_s \right) \\ \vdots & \ddots & \vdots \\ a_1 \exp\left(j2\pi L f_i/f_s \right) & \ldots & a_K \exp\left(j2\pi L f_i/f_s \right) \end{bmatrix} $$

Neglecting the receive noise, the received data $X = [x(1), x(2), \ldots, x(L)]$ can be expressed as

Fig. 1. monostatic L-shape MIMO radar
The expression in (9) could be discussed in detail in the next section.

III. IMPROVED PARAFAC ALGORITHM

A. Data Expansion Using Unitary Transformation

Employing the center of the ULA as the phase reference, the array manifold is conjugate centrosymmetric [23]. Given the receive array as an example, the central symmetry can be stated as $\Pi_{kk} a_{kk} (\theta, \phi_k) = a'_{kk} (\theta, \phi_k)$. Define a simple unitary matrix

$$Q_{2d} = \frac{1}{\sqrt{2}} \begin{bmatrix} I_d & jI_d \\ \Pi_d & -j\Pi_d \end{bmatrix} \in \mathbb{C}^{2d \times 2d}$$

if $d$ is even, or

$$Q_{2d+1} = \frac{1}{\sqrt{2}} \begin{bmatrix} I_d & 0 & jI_d \\ 0 & \sqrt{2} & 0 \\ \Pi_d & 0 & \Pi_d \end{bmatrix} \in \mathbb{C}^{(2d+1) \times (2d+1)}$$

when $d$ is odd. $Q_N$ is a sparse unitary matrix which maps $a_{kk} (\theta, \phi_k)$ onto an real-valued manifold. For example, if $N$ is odd, we have

$$a^o_k (\mu_k) = Q_N a_k (\mu_k) = \sqrt{2} \exp \left[ \frac{-(N-1)}{2} \pi \mu_k \right] \begin{bmatrix} \cos \left( \frac{N-1}{2} \pi \mu_k \right) \\ \cdots \\ \cos \left( \frac{1}{2} \pi \mu_k \right) \\ \sin \left( \frac{1}{2} \pi \mu_k \right) \\ \cdots \\ \sin \left( \frac{N-1}{2} \pi \mu_k \right) \end{bmatrix}^T$$

where $\mu_k = \sin \theta_k \cos \phi_k$. Obviously, the column vector $a^o_k (\mu_k)$ can be expressed as the product of a complex value and a real-valued vector. Let $Q = Q_N \otimes I_M$, we therefore have

$$Q^o a (\mu_k) = (Q_N^o a_k (\mu_k)) \otimes a (\mu_k) = a^o_k (\mu_k) \otimes a (\mu_k)$$

Define the transformed direction matrix $A^o_k$ as $A^o_k = Q_N^o A_k$. According to (6), we obtain $A^o_k = A^o_k \otimes A_t$ with $A^o_k = Q_N^o A_k$. The uncorrupted signals in (3) can be transformed into

$$X^o = Q^o X = Q^o A^o S^T = A^o S^T$$

On the other hand, due to the Vandermonde-like structure on $S$, we have $S^o \Pi_{kk} = A^o S^T$, where $A = \text{diag} (\beta_1, \beta_2, \ldots, \beta_k)$ is a diagonal matrix with the elements $\beta_k = \exp \{-j2\pi f_k (L+1) / f_k\}$, $(k = 1, \ldots, K)$. Notice that the source matrix $S$ has a similar central symmetry to $A_k$, another data matrix can be built

$$X_k = X_k \Pi_{kk} = A^o_k A_{kk} S^T = A^o_k S^T$$

one can easily find that $A_{kk} = [A^o_k \otimes A_t]$ with $A^o_k = A^o_k A^o$. Without additive noise, the expanded data matrix is formed as following

$$X_{kk} = [A^o_k \otimes A_{kk}] S^T = [A^o_k \otimes A_{kk}] S^T$$

where $A_{kk} \in \mathbb{C}^{2N \times K}$. The expression in (9) could also be counted as expanded PARAFAC decomposition in matrix format, which can be expressed in detail

$$X_{kk} = [A^o_k \otimes A_{kk}] S^T = \begin{bmatrix} X_1 \\ X_2 \\ \vdots \\ X_{2N} \end{bmatrix} = \begin{bmatrix} A^o_1 D_1 (A_{kk}) \\ A^o_2 D_2 (A_{kk}) \\ \vdots \\ A^o_{2N} D_{2N} (A_{kk}) \end{bmatrix} S^T$$

where $D_1 (A_{kk})$ denotes the operations of rearranging the $n$-th row of $A^o_{kk}$ into a diagonal matrix. The $n$-th slice $X_n$ of $X_{kk}$ is given by $X_n = A_{kk} D_n (A^o_{kk}) S^T$ ($n = 1, 2, \ldots, 2N$). Equation (10) shows that expanded PARAFAC model essentially doubles the number of available measurements from $MN$ to $2MN$. Consequently, increased estimation accuracy can be achieved by the expanded PARAFAC model. PARAFAC decomposition could be also expressed in scalar format

$$x_{m,n} = \sum_{l=1}^{L} A_{kk} (m, k) A^o_{kk} (n, k) s(l, k)$$

$(m = 1, 2, \ldots, M; n = 1, 2, \ldots, 2N; l = 1, 2, \ldots, L)$

The model in (11), which rearranged the data from the array antennas into a cube, can be regarded as the trilinear slice model. It depicts the echo form the perspective of three different diversities, as shown in Fig. 2. The matrix $X_n$ can be deem as slicing the cube data into successions of slices along the receive antenna direction. Due to the
symmetry of the parallel factor model, two more matrices can be constructed
\[
Y = \begin{bmatrix}
Y_1 \\
Y_2 \\
\vdots \\
Y_M \\
\end{bmatrix} = \begin{bmatrix}
SD_1(A_T) \\
SD_2(A_T) \\
\vdots \\
SD_M(A_T) \\
\end{bmatrix} A_T^T = [A_T \odot S] A_T^T \tag{12}
\]
and
\[
Z = \begin{bmatrix}
Z_1 \\
Z_2 \\
\vdots \\
Z_L \\
\end{bmatrix} = \begin{bmatrix}
A_E D_1(S) \\
A_E D_2(S) \\
\vdots \\
A_E D_L(S) \\
\end{bmatrix} A_E^T = [S \odot A_E] A_E^T. \tag{13}
\]
where the slices \(Y_m = SD_m(A_T) A_T^T (m = 1, 2, \ldots, M)\) and \(Z_l = A_E D_l(S) A_E^T (l = 1, 2, \ldots, L)\). The data matrices \(Y\) and \(Z\) can be interpreted as slicing the cube along the transmit antenna array direction the snapshot direction, respectively.

\[
\text{Fig. 2. Trilinear model}
\]

**B. Trilinear Alternate Least Squares (TALS)**

TALS is an effective algorithm to solve the PARAFAC model. The essence of TALS can be described as the following steps: (a) Fitting one of the slicing matrices \(X, Y\) or \(Z\) using Least Squares (LS) method, where the remaining two matrices are previously obtained. (b) Fitting the other two matrices in a similar way, and (c) Repeat (a) and (b) until LS cost function convergence. The details of solving the expanded PARAFAC model with TALS algorithm is described as follows.

Consider the situation where noise is present, according to (10), the LS fitting of \(X_{k}\) is
\[
f_X = \min_{A_y, A_S} \left\| \tilde{X} - [A_T \odot A_T] S^T \right\|_F
\]
where \(\tilde{X}\) represents the noisy \(X_{k}\). Therefore, the LS estimation for \(S\) is
\[
\hat{S}^T = \left[ \hat{A}_E \odot \hat{A}_T \right] \tilde{X}
\]
with \(\hat{A}_E\) and \(\hat{A}_T\) denote the estimated \(A_E\) and \(A_T\) in the last iteration. Similarly, the LS fitting of \(Y\) is
\[
f_Y = \min_{A_y, \hat{A}_E} \left\| \hat{Y} - [A_T \odot S] A_E^T \right\|_F
\]
where \(\hat{Y}\) denotes the noisy \(Y\). The LS update for \(A_E\) is
\[
\hat{A}_E^+ = [\hat{A}_T \odot \hat{S}] \hat{Y}
\]
where \(\hat{A}_T\) and \(\hat{S}\) are previously obtained. Finally, the LS fitting of \(Z\) is
\[
f_Z = \min_{A_y, \hat{A}_E} \left\| \hat{Z} - [S \odot A_E] A_E^T \right\|_F
\]
and the LS update for \(A_T\) is
\[
\hat{A}_T^+ = \left[ \hat{S} \odot \hat{A}_E \right] \hat{Z}
\]
C. **Angle Estimation**

The uniqueness property under mild conditions is a key feature of PARAFAC decomposition. Theorem 1 [24] gives the identifiability of the expanded PARAFAC model.

**Theorem 1**: \(X_{kn} = A_y D_n(A_E) S^T, (n = 1, 2, \ldots, 2N)\), where \(A_y \in \mathbb{C}^{m \times K}, A_E \in \mathbb{C}^{2N \times K}, S \in \mathbb{C}^{2 \times K}\). Consider that all the matrices are full \(k\)-rank [25], if the parameter identifiability satisfies
\[
k_{\lambda_y} + k_{\lambda_E} + k_S \geq 2K + 2
\]
then \(A_y, A_E\) and \(S\) are unique up to permutation and scaling of columns, where \(k_{\lambda_y}, k_{\lambda_E}\) and \(k_S\) denote the \(k\)-rank of \(A_y, A_E\) and \(S\), respectively. The estimated matrices \(\hat{A}_y, \hat{A}_E\) and \(\hat{S}\) satisfy \(\hat{A}_y = A_y \Pi \Lambda + N_1\), \(\hat{A}_E = A_E \Pi \Lambda + N_2\), and \(\hat{S} = S \Pi \Lambda + N_3\), where \(\Pi\) is a permutation matrix, \(N_1, N_2\) and \(N_3\) represent the corresponding estimation error, and \(\lambda_y, \lambda_E\) and \(\lambda_S\) stand for the diagonal scaling matrices satisfying \(\lambda_y \lambda_E \lambda_S = I_K\).

Once TALS is accomplished, the estimated direction matrices \(A_E\) and \(A_T\) could be obtained. Since the phase of the columns in \(A_T\) and \(A_E\) have linear characteristics, so the LS method is reused for 2D angle estimation. The estimated transmit vector \(a_t(\theta, \phi)\) was firstly normalized to the reference phase center. Let
\[
h_{t,1} = \text{angle}(\hat{a}_t(\theta, \phi))
\]
where the operation \(\text{angle}(\hat{a}_t(\theta, \phi))\) is to get the phase of \(\hat{a}_t(\theta, \phi)\). We construct the following matrix and vector
\[
P = \begin{bmatrix} 1 & 1 & \ldots & 1 \\ 0 & \pi & \ldots & (M-1)\pi \end{bmatrix} \in \mathbb{R}^{M \times 2}, \ u_t = \begin{bmatrix} u_{t,1} \\ u_{t,2} \end{bmatrix}
\]
Furthermore, the LS solution to \( u_k \) is
\[
\mathbf{u}_k = \mathbf{P}_k^\dagger \mathbf{h}_{k,1}
\]  
(21)

One can easily find that \( u_{k,2} \) is the LS estimation to \( \sin \theta_k \cos \phi_k \). Similarly, define
\[
\mathbf{h}_{k,2} = -\angle(\hat{\mathbf{a}}_k(\theta_k, \phi_k))
\]  
(22)

and construct
\[
\mathbf{P}_k = \begin{bmatrix}
0 & \ldots & 0 & 1 & \ldots & 1 \\
0 & \ldots & (N-1)\pi & 0 & \ldots & -(N-1)\pi \\
\end{bmatrix}
\]

Next, we got the LS solution to \( v_k \)
\[
\mathbf{v}_k = \mathbf{P}_k^\dagger \mathbf{h}_{k,2}
\]  
(24)

One can easily find that \( v_{k,2} \) is the LS estimation to \( \sin \theta_k \sin \phi_k \). The azimuth angle and elevation angle for the \( k \)-th target can be paired automatically through the following formula
\[
\begin{align*}
\theta_k &= \arcsin\left(\frac{u_{k,2} + v_{k,2}}{2}\right) \\
\phi_k &= \arctan\left(\frac{v_{k,2}}{u_{k,2}}\right)
\end{align*}
\]  
(25)

IV. ALGORITHM ANALYSIS

A. Identifiability

The formulation in (18) provides an upper bound on the identifiability of the proposed PARAFAC algorithm. Generally, we have \( k_{n_s} = M \), \( k_{n_s} = 2N \) and \( k_{n_s} = \min(K, L) \) in the trilinear model established in this study. When \( K \leq L \), the inequality in (18) can be rewritten as \( M + 2N - 2 \geq K \), which implies the maximum number of targets that the proposed algorithm can identify is \( M + 2N - 2 \), while the maximum number of targets that the traditional PARAFAC algorithm can detect is \( M + N - 2 \) [10]. Under the condition of \( K > L \), the proposed algorithm is effective when \( K \leq (M + 2N + L - 2)/2 \), while the conditional PARAFAC algorithm is effective when \( K \leq (M + N + L - 2)/2 \), therefore, the proposed algorithm could achieve better estimation accuracy under the same conditions.

B. Computation Complexity

Traditional Capon and MUSIC algorithm requires 2D peak searching in a higher dimensional space, which brings very heavy complexity. The computation complexity of the traditional ESPRIT algorithm is \( O\left(M^2N^2L + M^3 + N^3 + L^3\right) \), complexity in each TALS iteration of the traditional PARAFAC algorithm is \( O\left(MNLK + K^3\right) \), and the proposed MCS algorithm needs \( O\left(2MNLK + K^3\right) \) calculations per TALS iteration.

and algorithm convergence with only a few iterations. It is obvious that the computation complexity of the proposed algorithm is between the ESPRIT method and traditional PARAFAC algorithm.

C. Cramér-Rao Bound (CRB)

In this subsection, we derive CRB of 2D angle estimation in a monostatic MIMO radar. In particular, it provides an asymptotic lower bound on the 2D angle estimators. We assume that the signal \( \mathbf{s}(t) \) and the noise variance \( \sigma^2 \) are deterministic, and then estimation parameter vector is expressed as
\[
\mathbf{z} = [\theta_1, \ldots, \theta_K, \varphi_1, \ldots, \varphi_K]^T
\]  
(26)

Collect \( L \) snapshots and form the data matrix \( \mathbf{X} = [\mathbf{x}(1)^T, \mathbf{x}(2)^T, \ldots, \mathbf{x}(L)^T]^T \). The mean \( \mu \) and the covariance matrix \( \mathbf{G} \) of \( \mathbf{X} \) are
\[
\mu = \begin{bmatrix}
\mathbf{as}(1) \\
\vdots \\
\mathbf{as}(L)
\end{bmatrix}, \quad \mathbf{G} = \begin{bmatrix}
\sigma^2 \mathbf{I}_{MN} & 0 \\
0 & \sigma^2 \mathbf{I}_{MN}
\end{bmatrix}
\]  
(27)

From [26], we know that the \((i,j)\) element of the CRB matrix \( \mathbf{P}^{-1} \) can be expressed as
\[
\left[\mathbf{P}^{-1}\right]_{ij} = \text{tr}\left[\mathbf{H}^{-1} \mathbf{G} \mathbf{H}^{-1} \mathbf{G}^T\right] + 2 \text{Re}\left[\mu^{ij} \mathbf{G}^{-1} \mu^{ij}\right]
\]  
(28)

where \( \mathbf{G}^T \) and \( \mu^T \) are the derivative of \( \mathbf{G} \) and \( \mu \) on the \( i \)-th element of \( \mathbf{z} \), respectively. Since the covariance matrix is just related to \( \sigma^2 \), the first part of (28) can be ignored. Then
\[
\left[\mathbf{P}^{-1}\right]_{ij} = 2 \text{Re}\left[\mu^{ij} \mathbf{G}^{-1} \mu^{ij}\right]
\]  
(29)

Define \( \mathbf{a}_k = \mathbf{a}_k(\theta_k, \varphi_k) \otimes \mathbf{a}_k(\theta_k, \varphi_k) \), and we have
\[
\begin{align*}
\frac{\partial \mathbf{a}_k}{\partial \theta_k} &= \mathbf{a}_k(\theta_k, \varphi_k) \otimes \mathbf{a}_k(\theta_k, \varphi_k), \\
\frac{\partial \mathbf{a}_k}{\partial \varphi_k} &= \mathbf{a}_k(\theta_k, \varphi_k) \otimes \mathbf{a}_k(\theta_k, \varphi_k)
\end{align*}
\]  
(30)

Define
\[
\begin{align*}
\mathbf{d}_k &= \frac{\partial \mathbf{a}_k}{\partial \theta_k}, \\
\mathbf{f}_k &= \frac{\partial \mathbf{a}_k}{\partial \varphi_k}, \\
\mathbf{s}_k &= [s_k(1), \ldots, s_k(L)]^T
\end{align*}
\]  
(31)
with \( s_k(l) \) \((l=1, \ldots, L)\) denotes the \( k \)-th element of \( \mathbf{s}(l) \). Define the following matrix
Therefore we have
\[
\frac{\partial \mu}{\partial \zeta} = \Lambda, \quad \text{and}
\]
\[
2 \Re \left\{ \frac{\partial \mu}{\partial \zeta} \right\} = \frac{2}{\sigma^2} \Re \left\{ \Lambda^H \Lambda \right\}
\]
Thus, the CRB for the 2D angle estimation error in MIMO radar is
\[
\text{CRB} = \frac{\sigma^2}{2} \left( \Re \left\{ \text{D}^H \text{D} \right\} \right)^{-1}
\]
where \(\text{D} = [\text{D} \otimes \text{S}, \text{F} \otimes \text{S}]\), among which \(\text{D} = [d_1,...,d_k]\), \(\text{F} = [f_1,...,f_k]\).

V. SIMULATION RESULTS

To evaluate the performance of the proposed 2D angle estimation algorithm, 1000 Monte Carlo trials are taken. The signal-to-noise ratio in the simulation is defined by
\[
\text{SNR} = 10 \log_{10} \frac{\|	ext{X}_E\|}{\|	ext{X}_E - \hat{\text{X}}\|} \text{[dB]}
\]
The root mean squared error (RMSE) is used for performance assessment, which is defined as follows
\[
\text{RMSE} = \frac{1}{K} \sum_{k=1}^{K} \sum_{i=1}^{1000} \left[ (\hat{\theta}_k - \theta_i)^2 + (\hat{\phi}_k - \phi_i)^2 \right]
\]
where \(\hat{\theta}_k\) and \(\hat{\phi}_k\) denote the estimation of elevation and azimuth angle of the \(k\)-th target in the \(i\)-th Monte Carlo trial. In the simulation, \(M\), \(N\) and \(L\) denote the number of the transmit elements, receive elements and snapshots. We consider that there are \(K = 3\) targets located at angles \((\theta_1, \phi_1) = (10, 15)\), \((\theta_2, \phi_2) = (20, 25)\) and \((\theta_3, \phi_3) = (30, 35)\).

Fig. 3 present the 2D-DOA estimation results of our algorithm with \(\text{SNR} = 5\text{dB}\). Parameters are set to \(M = N = 8\) and \(L = 40\), respectively. As seen in this figure, our algorithm is able to estimate azimuth angle and elevation angle for monostatic L-shape MIMO radar.

Fig. 4 depicts the performance comparison of our algorithm, the conventional ESPRIT algorithm, the traditional PARAFAC algorithm and the CRB with \(M = N = 8\) and \(L = 40\). According to Fig. 4, the Unitary-ESPRIT algorithm has estimation performance very close to that of the ESPRIT method. Besides, all algorithms would achieve better RMSE performance with the growing SNR, and our algorithm outperform the other algorithms with the same SNR, especially in the condition of low SNR. Fig. 5 illustrate that the RMSE comparison of the above algorithms with different snapshots number with \(M = N = 8\) and \(\text{SNR}=0\text{dB}\). All algorithms exhibit an increasing RMSE trend with the increase in the number of snapshots, and it is obvious that our algorithm has much better 2D angle estimation than ESPRIT and the traditional PARAFAC algorithm.

Fig. 6. RMSE performance with different \(M\)

The RMSE performance of our algorithm under the condition of different transmit elements number \(M\) and different receive antennas number \(N\) are investigated separately in the simulation, as shown in Fig. 6 and Fig. 7. \(N = 8\), \(L = 50\) and \(M = 8\), \(L = 50\) are considered in Fig. 6.
and Fig. 7, respectively. By comparing the above results, one can clearly see that the 2D angle estimation performance of our algorithm is gradually improving with the increasing antenna number. This is commensurate with MIMO systems, as estimation performance can be improved with multiple antennas because of diversity gain.

Fig. 9. RMSE performance with different \( K \).

Fig. 7. RMSE performance with different \( N \).

VI. CONCLUSION

An expanded PARAFAC algorithm is developed for 2D angle estimation in monostatic L shape MIMO radar. The virtual aperture of the proposed algorithm is expand with unitary transform, and 2D angle estimation problem in our work was linked to the PARAFAC model. Our algorithm is attractive from the perspective estimation accuracy. It achieves better angle estimation performance to the traditional ESPRIT method and PARAFAC algorithm. Our algorithm doesn’t require singular value decomposition of the received data while automatically paired the estimated angles, which means it have blind and robust characteristic. Also, simulation results showed that improvement of the proposed algorithm.

REFERENCES


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