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Joint Estimation of DOA and TDOA of Multiple Reflections in Mobile Communications

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ABSTRACT In a multipath communication scenario, it is often relevant to estimate the directions and relative delays of each multipath signal. We present an effect algorithm for the simultaneous estimation of these parameters by re-iterative super-resolution, beamforming, and MUSIC-like searching techniques. The algorithm first separates and estimates direction of arrival (DOA) of the multipath signals. Linear constrained minimum power beamforming is used to obtain the transmitted time function of the desired signal in certain incident angle. Then, time difference of arrival (TDOA) for the incident signals are mapped into phase shifts in the frequency domain. Using the DFT of the desired signal and the received data, we can separate the phase shifts in the frequency domain due to time delay by MUSIC-like searching. At the same time, the pairing of the estimated DOA and TDOA is automatically determined. Computer simulations illustrating the performance of the proposed algorithm with the Cramer–Rao bound are included.

INDEX TERMS Wireless communication, array signal processing, joint TDOA-DOA estimation, multipath signal.

I. INTRODUCTION

With the increasing demand for the mobile communication system, innovative approaches are needed to improve the performance for overcoming errors caused by mobile channel. Long Term Evolution (LTE) is considered as one of the promising systems that is suitable to these requirements [1], [2]. As the signal from the User Equipment (UE) to base station undergoes multipath rays of the direct signal, the multipath signals with different DOAs and time delays combined with direct signal will degrade and influence the estimation of the desired signal. Channel estimation of the base station involves the DOAs and time delays parameter estimation for multipath and direct sources transmitted by the UE. So joint DOA and TDOA estimation by the base station is of interest for advanced handover schemes, emergency localization, and potentially many user services [3].

Many space-time processing approaches have been investigated to estimate the DOA and TDOA with the smart array for obtaining the desired signals. In [4] and [5], the approaches require to ML searches with great computation load or need good accurate initial value for the accurate estimation. In [6]–[8], the algorithms are developed to transform the data in frequency domain and maps delays into phase shifts, and the joint estimation problem is changed to one that can be solved using 2-D ESPRIT techniques. Joint DOA and TDOA estimation by multi-invariance MUSIC are proposed in [9] and [10]. Those algorithms require the known modulation pulse shape of one symbol to recovery the time delays. And the delay less than the symbol duration can be obtained by the algorithms. In mobile communication, time delays between different paths are most likely to be longer than the symbol rate. The algorithms [11], [12] do not need the known modulation pulse shape instead of the known transmitted signal. The algorithms require the preamble to recovery the time delays between each arrival. In [13], a method is proposed, which associates DOA from the MUSIC algorithm and TDOA from the correlator separatively. It considered multipath signals as independents sources, assumed the delays between the components are larger than the duration of autocorrelation of emitted signals.

In this work, a practical method to estimate TDOA associated with DOA information in the presence of multipath without any preamble sequence is proposed. The algorithm is based on an efficient high-resolution scheme that transforms the multi-dimensional estimation involved in two sets of simple one-dimensional (1-D) estimation. To do so, we propose to use the RISR algorithm to estimate DOAs of the multipath signals without forward-backward averaging and spatial smoothing [14]. And the LCMP beamforming [15] in conjunction with the received data by the array is used to estimate the time function of the desired signal in certain incident angle. Then time delays of the incident signals are mapped into phase shifts in frequency domain by Discrete Fourier Transformation (DFT). Using the sequence of the desired signal and the received data in frequency domain, we can estimate the time delays of the signals by MUSIC-like peaks searching. The new proposed numerical method is able to associate correctly the DOA from the RISR algorithm and TDOA from the MUSIC-like algorithm.

Our major contributions are summary as follows: (1) Use the RISR, beamforming and MUSIC-like techniques in conjunction to realize the DOA and TODA estimation without preamble knowledge (2) The algorithms is suitable of arbitrary array structures (3) Time delay within a fraction of the sampling time can be estimated accurately in frequency domain. To our knowledge, no other existing algorithm is able to jointly estimate of TDOA and DOA of the multipath signals without preamble.

The structure of the paper is as follows. In Section II, we begin our discussion by formulating the problem and describing the data model. Section III contains a detailed derivation of the basic steps of the algorithm, including the techniques to obtain the DOAs of multiple sources and to estimate the time function of them. Identifiability of the DOAs and TDOAs using the proposed technique is addressed in Section IV. Section V illustrates the performance using computer simulations.

Notation: Vectors (matrices) are denoted by boldface lower (upper) case letters, all vectors are column vectors, superscripts $(\cdot)^T$, $(\cdot)^*$, $(\cdot)^H$ and $(\cdot)^\dagger$ denote transpose, conjugate, complex conjugate transpose and Pseudo-inverse respectively, $E[\cdot]$ denotes statistical expectation, $\boldsymbol{I}$ is the identity matrix, $|\cdot|$ represents the modulus of a complex number, and $\|\cdot\|$ is the Euclidean norm of a vector. The symbol $\odot$ denotes the Hadamard product between two matrices of appropriate size and the symbol $\circ$ is used to denote convolution operator. product.

II. SYSTEM MODEL

Consider a linear array composed of M omnidirectional sensors in a base station to receive $K(K < M)$ narrow-band plane wave signals from directions $\theta_1, \theta_2, \cdots \theta_K$, in which one of the signals is the direct signal and the other $K-1$ signals are reflection components. The distance between the first reference element and the i-th element is d_i in the array with at least two elements spaced at half of the wavelength of radiation sources or closer, as shown in Fig. 1. Signal $s_d(t)$ is defined as the direct signal with fading coefficient $\beta_d = 1$ and time delay $\tau_d = 0$, then the other reflection signals received can be expressed as

$$s_k(t) = \beta_k s_d(t - \tau_k). \tag{1}$$

In general, array antenna received signal vector $\boldsymbol{y}(t)$ formed by the superposition of the multipath signals and noise can be described as

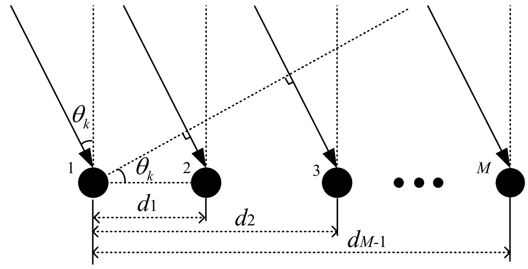

FIGURE 1. Schematic diagram of an antenna array with one impinging signal.

$$\boldsymbol{y}(t) = \sum_{k=1}^{K} \boldsymbol{a}(\theta_k) s_k(t) + \boldsymbol{v}(t) \tag{2}$$

where $\boldsymbol{y}(t) = [y_1(t), y_2(t), \cdots, y_M(t)]^T$ is the M dimensional snapshot data vector of the array, $\boldsymbol{v}(t) = [v_1(t), v_2(t), \cdots, v_M(t)]^T$ is the M dimensional noise data vector. The vector-valued function $\boldsymbol{a}(\theta_k)$ is the array response vector (steering vector) for an array of M elements to the kth source signal from the direction θ_k, expressed as follows:

$$\boldsymbol{a}(\theta_k) = [1, e^{-j\phi_{k1}}, \cdots, e^{-j\phi_{ki}}, \cdots, e^{-j\phi_{k(M-1)}}]^T \tag{3}$$

where the phase shift of the ith element for each narrowband arrival signal can be defined as $\phi_{ki} = 2\pi f_c d_i \sin(\theta_k) \big/ c$. f_c is the carrier frequency of the incident signals, c is the speed of light. It is assumed that the signals and noise are stationary, zero mean uncorrelated random processes. Further, the noise vector is Additive Gaussian White Noise (AGWN) with variance σ^2.

III. REVIEW OF SOME TECHNIQUES

The proposed algorithm is partly based on the RISR algorithm and beamforming techniques, so it is necessary to make a brief review of them.

A. RISR DOA ESTIMATION ALGORITHM

Many traditional DOA estimation methods are based on the covariance information of spatial samples, and the estimation performance will be significantly decreased when some sources are related in time domain. The essence of Re-Iterative Super-Resolution (RISR) algorithm is the recursive implementation of the minimum mean-square error (MMSE) [16]. When the determined spatial covariance information and the approximate value of the array calibration tolerance are given, RISR determines an MMSE filter bank according to the estimate of the spatial power distribution. Then the MMSE filter bank is applied to the received data for the updated estimation of spatial power distribution which is used to redefine the MMSE filter banks. The two processes above constantly alternates. Finally, the number of signal sources (the prior knowledge of the number of signal source is not needed), their respective DOAs, and their respective magnitudes can be determined automatically.

One of the advantages for the algorithm is that it can distinguish the coherent signals received by the array, which often happen in the mobile communication.

Given K narrowband signals from the far field which simultaneously incident on the array. After A/D conversion, The lth time sample, which contains superposition of signal and noise can be expressed in vector notation, same as (2)

$$\boldsymbol{y}(l) = \sum_{k=1}^{K} \boldsymbol{x}_k(l) + \boldsymbol{v}(l) \tag{4}$$

where $\boldsymbol{x}_k(l) = \boldsymbol{a}(\theta_k)s_k(l)$. For simply addressing DOA estimation, let $\psi_k = 2\pi f_c \sin\theta_k / c$. Then the steering vector $\boldsymbol{a}(\theta_k)$ in θ-space is transformed to ψ-space as $\boldsymbol{a}(\psi_k)$. In order to determine the DOA of each signal in ψ-space at the lth sample, we approximate received data $\boldsymbol{y}(l)$ in (4) with a parameterized version as

$$\boldsymbol{y}(l) \cong \tilde{\boldsymbol{y}}(l) \triangleq \boldsymbol{A}\boldsymbol{s}(l) + \boldsymbol{v}(l). \tag{5}$$

According to the array configure, the $M \times N$ dimensional array manifold matrix $\boldsymbol{A}$ is constructed and defined as

$$\begin{aligned} \boldsymbol{A} &= [\boldsymbol{a}(0), \boldsymbol{a}(\psi_\Delta), \cdots, \boldsymbol{a}((N-1)\psi_\Delta)] \\ &= \begin{bmatrix} 1 & 1 & \cdots & 1 \\ 1 & e^{jd_1\psi_\Delta} & \cdots & e^{jd_1\psi_\Delta(N-1)} \\ \vdots & \vdots & \cdots & \vdots \\ 1 & e^{jd_{M-1}\psi_\Delta} & \cdots & e^{jd_{M-1}\psi_\Delta(N-1)} \end{bmatrix} \end{aligned} \tag{6}$$

which consists of N spatial steering vectors corresponding to angles defined over 2π with the angular increment $\psi_\Delta = 2\pi/N$. The angle range of incident signals is separated to N grids. With the parameter N increasing, the quantization effect of the angle becomes better, while the capability to solve the problem of noise, source correlation effects and array modeling errors becomes stronger.

The idea of the RISR algorithm is to estimate the Minimum Mean-Square Error through the $M \times N$ adaptive filter bank, with the received signal model defined in (5) and the assumed knowledge of spatial complex vector $\boldsymbol{s}(l)$. The MMSE cost function can be written as

$$J(\boldsymbol{W}) = \left\| \boldsymbol{s}(l) - \boldsymbol{W}^H(l)\boldsymbol{y}(l) \right\|^2. \tag{7}$$

Minimizing (7) and the well-known MMSE filter can be obtained as

$$\boldsymbol{W}(l) = (E\{\boldsymbol{y}(l)\boldsymbol{y}^H(l)\})^{-1}E\{\boldsymbol{y}(l)\boldsymbol{s}^H(l)\}. \tag{8}$$

Substituting $\boldsymbol{y}(l)$ of (5) into (8), with the added assumption that the signal and noise are uncorrelated and statistically independent, then results is found to be

$$\boldsymbol{W}(l) = (\boldsymbol{A}E\{\boldsymbol{s}(l)\boldsymbol{s}^H(l)\}\boldsymbol{A}^H + E\{\boldsymbol{v}(l)\boldsymbol{v}^H(l)\})^{-1}\boldsymbol{A}E\{\boldsymbol{s}(l)\boldsymbol{s}^H(l)\} \tag{9}$$

where noise covariance matrix $\boldsymbol{R}_v = E\{\boldsymbol{v}(l)\boldsymbol{v}^H(l)\}$, can be written in the form as $\sigma^2\boldsymbol{I}$. As a measure of statistical similarity over time, the temporal correlation of the signal sources is meaningless in terms of the RISR algorithm because the RISR formulation just works on each snapshot independently or combines power estimates via non-coherent integration. In order to perform the temporally uncorrelated assumptions, the spatial energy distribution matrix can be written as

$$\boldsymbol{P}(l) = E\{\boldsymbol{s}(l)\boldsymbol{s}^H(l)\} \odot \boldsymbol{I}_M \tag{10}$$

where $\boldsymbol{I}_M$ is an $M \times M$ identity matrix. The diagonal elements of $\boldsymbol{P}(l)$ comprise the power distribution in space. Substituting the noise covariance matrix and spatial power distribution matrix of (10) into (9) yield

$$\boldsymbol{W}(l) = (\boldsymbol{A}\boldsymbol{P}(l)\boldsymbol{A}^H + \boldsymbol{R}_v)^{-1}\boldsymbol{A}\boldsymbol{P}(l). \tag{11}$$

Given the filter bank $\boldsymbol{W}(l)$, the MMSE estimation of $\boldsymbol{s}(l)$ can be obtained as

$$\hat{\boldsymbol{s}}(l) = \boldsymbol{W}^H(l)\boldsymbol{y}(l) \tag{12}$$

Since the prior knowledge of $\boldsymbol{P}(l)$ is unknown, the recursive process of the previous estimation is needed. Based on the matched filter bank strategy, the pre-estimate of $\boldsymbol{s}(l)$ can be determined as

$$\hat{\boldsymbol{s}}_0(l) = \boldsymbol{A}^H\boldsymbol{y}(l). \tag{13}$$

So the initial estimation $\boldsymbol{P}(l)$ subsequently is computed as

$$\hat{\boldsymbol{P}}_0(l) = [\hat{\boldsymbol{s}}_0(l)\hat{\boldsymbol{s}}_0^H(l)] \odot \boldsymbol{I}_M. \tag{14}$$

The relation between new MMSE filter bank estimation $\hat{\boldsymbol{W}}_i(l)$ and previous estimation $\hat{\boldsymbol{P}}_{i-1}(l)$ is

$$\hat{\boldsymbol{W}}_i(l) = (\boldsymbol{A}\hat{\boldsymbol{P}}_{i-1}(l)\boldsymbol{A}^H + \boldsymbol{R}_v)^{-1}\boldsymbol{A}\hat{\boldsymbol{P}}_{i-1}(l), \tag{15}$$

then a new MMSE estimation of $\boldsymbol{s}(l)$ can be expressed as follow

$$\hat{\boldsymbol{s}}_i(l) = \hat{\boldsymbol{W}}_i^H(l)\boldsymbol{y}(l). \tag{16}$$

The ith spatial power distribution estimation has the same expression form like (14)

$$\hat{\boldsymbol{P}}_i(l) = [\hat{\boldsymbol{s}}_i(l)\hat{\boldsymbol{s}}_i^H(l)] \odot \boldsymbol{I}_M. \tag{17}$$

Recursive procedure for spatial power distribution estimation may be stopped until $\left\| \hat{\boldsymbol{s}}_i(l) - \hat{\boldsymbol{s}}_{i-1}(l) \right\|^2 < \varepsilon$, where the ε is a pre-specified threshold or after some pre-determined number of recursions. At the end of the recursion, the diagonal elements of the diagonal matrix $\sqrt{\boldsymbol{P}_i(l)}$ is "RISR spectrum" and the spatial magnitude distribution consistent with the ψ-space division for the radiation sources is given. After that, we can calculate the DOAs of the signals recieved by

$$\theta_k = arc\sin\frac{\psi_k c}{2\pi f_c}. \tag{18}$$

B. LCMP BEAMFORMING

Weighted vector $\boldsymbol{w}$ of LCMP (linear constrained minimum power) beamforming algorithm [15] can be obtained by minimizing the output power, under M_c linear constraint conditions as

$$\boldsymbol{w}^H \boldsymbol{C} = \boldsymbol{g}^H \tag{19}$$

where $\boldsymbol{w}$ is an M-dimensional vector, $\boldsymbol{g}^H$ is an M_c-dimensional vector and $\boldsymbol{C}$ is an $M \times M_c$ matrix, the column vectors of which are linearly independent. Assumed that the first column of $\boldsymbol{C}$ corresponds to the steering vector of the signal of interest, and the first element of $\boldsymbol{g}$ is unit, the signal will pass through the filter undistorted. Furthermore the data covariance matrix $\boldsymbol{R}_y$, which keeps the output power minimum in the constraint condition of (19) is known as [15]

$$P_{out} = \boldsymbol{w}^H \boldsymbol{R}_y \boldsymbol{w} \tag{20}$$

minimize the following function

$$J \triangleq \boldsymbol{w}^H \boldsymbol{R}_y \boldsymbol{w} + [\boldsymbol{w}^H \boldsymbol{C} - \boldsymbol{g}^H]\lambda + \lambda^H [\boldsymbol{C}^H \boldsymbol{w} - \boldsymbol{g}], \tag{21}$$

we can obtain

$$\lambda^H = -\boldsymbol{g}^H [\boldsymbol{C}^H \boldsymbol{R}_y^{-1} \boldsymbol{C}]^{-1} \tag{22}$$

and

$$\boldsymbol{w}_{\mathrm{LCMP}}^H = \boldsymbol{g}^H [\boldsymbol{C}^H \boldsymbol{R}_y^{-1} \boldsymbol{C}]^{-1} \boldsymbol{C}^H \boldsymbol{R}_y^{-1}. \tag{23}$$

In our case, we will use LCMP beamforming to estimate the time function of one desired signal with other signals suppressed. According to (19), the distortionless constraint with adding constraints, can be written as:

$$\begin{cases} \mathbf{w}^H \mathbf{a}(\theta_k) = 1 \\ \mathbf{w}^H \mathbf{a}(\theta_i) = 0, \quad i \in [1, \ldots, K] \text{ and } i \neq k \end{cases} \tag{24}$$

IV. PROPOSED ALGORITHM

Our goal is to jointly estimate DOA and TDOA parameters without knowledge of the emitter signal under multipath propagation conditions. Firstly, the RISR algorithm presented in III-A will be applied to separate the coherent signals and determine DOAs of them. Furthermore, the LCMP beamforming is established according to the DOA information obtained. The optimal beamformer can maximize the ratio of interest output signal to the sum power of interference (other signals) and noise. It is also considered as a spatial filter, which makes the useful signal pass and make the output power of the noise and interference as small as possible. Next, we will find the delays of different impinging signals in the array. MUSIC-like searching method is used to estimate the delays through the orthogonality principle between the signal subspace and the noise subspace. There are different peaks in the searching which are related to the delays of arrival, associated each delay with its proper DOA.

A. MUSIC-LIKE ALGORITHM FOR TIME DELAY ESTIMATION

MUSIC is explored into high-resolution time delay estimation though the conventional MUSIC algorithm [17] was proposed to estimate the DOAs of the sources. As the data model description in (1) and (2), the wireless multipath channel containing K path components can be modeled by

$$\boldsymbol{h}(t) = \sum_{k=1}^{K} \boldsymbol{a}(\theta_k)\beta_k \delta(t - \tau_k) \tag{25}$$

where $\delta(t)$ is the dirac pulse. Therefore, the signal model can be rewritten as

$$\boldsymbol{y}(t) = s(t) \circ \boldsymbol{h}(t) + \boldsymbol{v}(t) \tag{26}$$

$$= \sum_{k=1}^{K} \boldsymbol{a}(\theta_k)\beta_k s(t - \tau_k) + \boldsymbol{v}(t). \tag{27}$$

The DFT expression of $\boldsymbol{y}(t)$ in (26) can be written as

$$\begin{aligned} \boldsymbol{Y}(f) &= S(f)\boldsymbol{H}(f) + \boldsymbol{V}(f) \\ &= S(f) \sum_{k=1}^{K} \boldsymbol{a}(\theta_k)\beta_k e^{-j2\pi f \tau_k} + \boldsymbol{V}(f) \\ &= [\boldsymbol{A}_\theta \boldsymbol{\beta}]\,[S(f)\boldsymbol{b}_\tau] + \boldsymbol{V}(f) \end{aligned} \tag{28}$$

where $S(f)$, $\boldsymbol{H}(f)$ and $\boldsymbol{V}(f)$ are the Fourier transform of $s(t)$, $\boldsymbol{h}(t)$ and $\boldsymbol{v}(t)$, respectively. $\boldsymbol{A}_\theta = [\boldsymbol{a}(\theta_1), \boldsymbol{a}(\theta_2), \cdots, \boldsymbol{a}(\theta_K)]$, $\boldsymbol{\beta} = \mathrm{diag}\{[\beta_1, \beta_2, \cdots, \beta_K]\}$ and $\boldsymbol{b}_\tau = [e^{j2\pi f \tau_1}, \cdots, e^{j2\pi f \tau_K}]^T$. The diag$\{\cdot\}$ is the diagonal operator, which form a vector creates a diagonal matrix, whose elements are the elements of the vector. As stated in [18], we transform the data received into frequency domain and construct an $M \times L$ matrix

$$\begin{aligned} \boldsymbol{Y}_L &= [\boldsymbol{Y}(f_1), \boldsymbol{Y}(f_2), \cdots, \boldsymbol{Y}(f_L)] \\ &= [\boldsymbol{A}_\theta \boldsymbol{\beta}]\left[\boldsymbol{B}_\tau \mathrm{diag}\{\boldsymbol{S}(f)\}\right] + \boldsymbol{V}_L \\ &= \boldsymbol{A}_{[\beta,\theta]} \boldsymbol{B}_{[f,\tau]}^T + \boldsymbol{V}_L \end{aligned} \tag{29}$$

where

- L represents the number of different frequency points in frequency domain
- $\boldsymbol{A}_{[\beta,\theta]} = \boldsymbol{A}_\theta \boldsymbol{\beta}$ is an $M \times K$ matrix
- $\boldsymbol{S}(f) = [S(f_1), S(f_2), \cdots, S(f_L)]$ is the L-point DFT of the signal function
- $\boldsymbol{V}_L$ is an $M \times L$ matrix, which presents the noise in frequency domain.
- $\boldsymbol{B}_\tau$ is a $K \times L$ matrix, which can be written as

$$\boldsymbol{B}_\tau = \begin{bmatrix} e^{-j2\pi f_1 \tau_1} & e^{-j2\pi f_2 \tau_1} & \cdots & e^{-j2\pi f_L \tau_1} \\ e^{-j2\pi f_1 \tau_2} & e^{-j2\pi f_2 \tau_2} & \cdots & e^{-j2\pi f_L \tau_2} \\ \vdots & \vdots & \ddots & \vdots \\ e^{-j2\pi f_1 \tau_K} & e^{-j2\pi f_2 \tau_k} & \cdots & e^{-j2\pi f_L \tau_k} \end{bmatrix} \tag{30}$$

- $\boldsymbol{B}_{[f,\tau]} = [\boldsymbol{B}_\tau \mathrm{diag}\{\boldsymbol{S}(f)\}]^T$ is a $L \times K$ matrix, the ith column of which can be expressed as

$$\boldsymbol{b}_i = [S(f_1)e^{-j2\pi f_1 \tau_i}, \cdots, S(f_L)e^{-j2\pi f_L \tau_i}]^T. \tag{31}$$

As the analysis above, the transpose of $\boldsymbol{Y}_L$ can be expressed as

$$\boldsymbol{Y}_L^T = \boldsymbol{B}_{[f,\tau]}\boldsymbol{A}_{[\beta,\theta]}^T + \boldsymbol{V}_L^T = [\boldsymbol{b}_1, \boldsymbol{b}_2, \cdots, \boldsymbol{b}_K]\begin{bmatrix} \beta_1 \boldsymbol{a}^T(\theta_1) \\ \beta_2 \boldsymbol{a}^T(\theta_2) \\ \vdots \\ \beta_K \boldsymbol{a}^T(\theta_K) \end{bmatrix} + \boldsymbol{V}_L^T. \quad (32)$$

Comparing it with (2), $\boldsymbol{Y}_L^T$ in (32) behaves like the received signal at an array whose manifold matrix is $\boldsymbol{B}_{[f,\tau]}$ and the equivalent signal matrix is represented by $\boldsymbol{A}_{[\beta,\theta]}^T$. The columns of $\boldsymbol{B}_{[f,\tau]}$ behave like the steering vectors in an array with L virtual sensors. The correlation matrix of $\boldsymbol{Y}_L^T$ can be written as

$$\begin{aligned} \boldsymbol{R}_{YY} &= E[\boldsymbol{Y}_L^T\left(\boldsymbol{Y}_L^T\right)^H] \\ &= \boldsymbol{B}_{[f,\tau]}E[\boldsymbol{A}_{[\beta,\theta]}^T(\boldsymbol{A}_{[\beta,\theta]}^T)^H]\boldsymbol{B}_{[f,\tau]}^H + E[\boldsymbol{V}_L^T\left(\boldsymbol{V}_L^T\right)^H] \\ &= \boldsymbol{B}_{[f,\tau]}\boldsymbol{P}_{[\beta,\theta]}\boldsymbol{B}_{[f,\tau]}^H + \boldsymbol{R}_{VV} \end{aligned} \quad (33)$$

where $\boldsymbol{P}_{[\beta,\theta]} = E[\boldsymbol{A}_{[\beta,\theta]}^T(\boldsymbol{A}_{[\beta,\theta]}^T)^H]$ is the covariance matrix with rank K. Assumed $L > K$, the rank of matrix $\boldsymbol{R}_{YY}$ will be K without noise effect. We can apply EVD(eigenvalue decomposition) to $\boldsymbol{R}_{YY}$

$$\boldsymbol{R}_{YY} = \boldsymbol{U}_S\boldsymbol{\Sigma}_S\boldsymbol{U}_S^H + \boldsymbol{U}_N\boldsymbol{\Sigma}_N\boldsymbol{U}_N^H \quad (34)$$

where $\boldsymbol{\Sigma}_S$ stands for a $K \times K$ diagonal matrix whose diagonal elements are the largest K eigenvalues, corresponding to $\boldsymbol{U}_S = [\boldsymbol{u}_1 \cdots \boldsymbol{u}_K]$, called signal subspace. And $\boldsymbol{\Sigma}_N$ stands for a diagonal matrix whose diagonal elements are composed of the smallest $L-K$ eigenvalues, corresponding to $\boldsymbol{U}_N = [\boldsymbol{u}_{K+1} \cdots \boldsymbol{u}_L]$, called noise subspace. By utilizing the orthogonality principle between the signal subspace and the noise subspace, we can get the time delay estimation of the sources in frequency domain by MUSIC-like searching as

$$F(\tau) = \frac{1}{\boldsymbol{b}^H\boldsymbol{U}_N\boldsymbol{U}_N^H\boldsymbol{b}} \quad (35)$$

where $\boldsymbol{b} = [S(f_1)e^{-j2\pi f_1\tau}, \cdots, S(f_L)e^{-j2\pi f_L\tau}]^T$ has same form as (31). The estimate of multipath time delay τ is

$$\hat{\tau} = \arg\max_{\tau} F \quad (36)$$

Searching the time delay variable τ with a constant increment in the delay range by the MUSIC-like method, we finally obtain TDOA of the signals.

B. REALIZATION OF JOINT DOA AND TDOA ESTIMATION BY THE PROPOSED ALGORITHM

The algorithm proposed in this paper has many advantages in DOA estimation, RISR algorithm is applicable to arbitrary array as long as the array manifold is known. And compared with other convectional algorithms [17], [19], RISR can estimate the direction of arrival accurately in the case that the number of multipath sources is unknown. In addition, many conventional time delay estimation algorithms take advantage of the cross-relation between the transmitted signal and the received signal while Rayleigh restriction limits the resolution of these algorithms. The MUSIC-like time delay estimation algorithm has more high resolution for delay estimation, which can distinguish the multipath signal components of the time intervals below the Rayleigh limit.

The specific steps are as follows:

1) Calculate the angles of incidence signals by RISR algorithm, and pay attention to select the appropriate number of iterations. According to the observation, RISR can always reach steady-state less than 15 iterations, regardless of the number of signal source or the spatial structure of the array.
2) Apply Beamforming technique to one of the interest directions obtained in step 1), by phasing the array to steer the main lobe in the specific direction θ_k so that all the received signals except the specific direction can be eliminated. In narrowband beamforming, a complex weight is applied to the signal at each sensor and summed to form the beamformer output
$$s_k(t) = \boldsymbol{w}^H(\theta_k)\boldsymbol{y}(t). \quad (37)$$
We make $\boldsymbol{w} = \boldsymbol{w}_{\mathrm{LCMP}}$ which is calculated as (23), the $s_k(t)$ in the direction θ_k will be well recovered .
3) MUSIC-like algorithm is applied to the time delay estimation. As described in IV-A. The arbitrary signal $s_p(t), p \in [1, \cdots, K]$, obtained from the previous step of the beamforming can be assumed as a reference signal(maybe not the the direct signal). Then we transform the reference signal to frequency domain, we can obtain $S_p(f) = S(f)e^{-j2\pi f\tau_p}$. And construct the steering vector of time delay τ
$$\boldsymbol{b} = [S_p(f_1)e^{-j2\pi f_1\tau}, \cdots, S_p(f_L)e^{-j2\pi f_L\tau}]^T \quad (38)$$
Substituting of (38) into (35), use the spectral searching with respect to τ in the specified range of time delay, we can get K peaks, which correspond to the relative delay to the signal $S_p(f)$. Since $s_p(t) = \beta_p s_d(t-\tau_p)$ according to (1), so MUSIC-like algorithm will produce spectral peaks on $\tau_k = \tau + \tau_p$ in the specific search range, which is relative time delay to τ_p. The real time delay of K path signals would be
$$\boldsymbol{\tau} = [\tau_1 - \tau_p, \cdots, \tau_K - \tau_p]. \quad (39)$$
In fact, we always believe that the arrival time τ_d of the direct signal on the antenna is the shortest. Assumed that $\tau_d = 0$, $\tau_d - \tau_p$ is the minimum element in $\boldsymbol{\tau}$. Thus, we have
$$d = \arg\min_k[\tau_k - \tau_p]. \quad (40)$$
Finally, the real delay time of each arrival signal to the direct signal can be obtained by the calculation as follow
$$\boldsymbol{\tau} - (\tau_d - \tau_p). \quad (41)$$

If we want to link DOA with the right TDOA, signals in other arrival directions provided by the beamformer are considered as reference signals to estimate relative time delays by the MUSIC-like method, respectively. According to the relations of the positions of spectral peaks in multiple searching, the DOA and TDOA can be paired correctly.

V. SIMULATION RESULTS

In this section, we conduct several simulations to demonstrate the performance of the proposed algorithm. The incident narrow-band multipath signals with carrier frequency $f_c = 300$MHz and sampling rate f_s =1GHz under the AWGN background are received by a 10-element uniform linear array (ULA), in which the distance between adjacent elements is half-wavelength. And the number of snapshots is 100.

(a) Suppose number of multipaths are $K = 3$ and incident directions of three related signals are $\theta = [-30°, 15°, 60°]$ corresponding to time delays $\boldsymbol{\tau} = [3.2,\ 0,\ 6.7]ns$. And the fading coefficient are $\beta = [0.4e^{j\frac{1}{3}\pi},\ 1,\ 0.7e^{j\frac{1}{5}\pi}]$. We consider that the direct signal comes from the direction of $\theta = 15°$. Under the condition of SNR=20dB, RISR algorithm is applied to estimate DOA with phase angle ψ quantization $N = 720$, see (6). After iterating 10 times, the pseudo-spectrum for DOA estimation is shown in Fig.2. Three spectral peaks of the spatial energy distribution by RISR are very sharp, corresponding to three estimation values $[-29.82°, 14.97°, 60.39°]$.

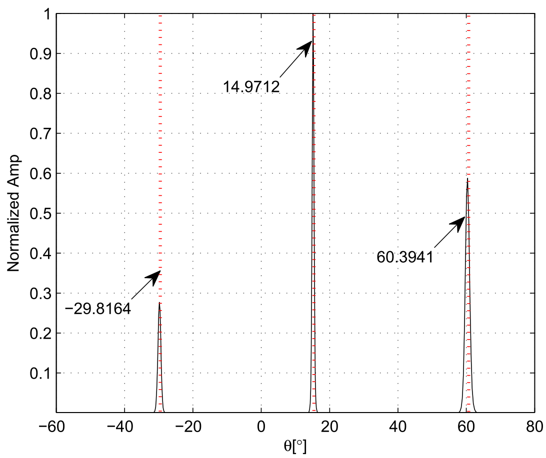

FIGURE 2. DOA estimation in 20dB with RISR pseudospectrum.

After getting incident angles of the signals, we need to estimate the time function of signals using the beamforming before realizing the high-resolution time delay estimation of each arrival signal. We firstly choose the signal from the direction $\hat{\theta}_p = -29.82°$ as the reference signal, and use LCMP beamforming in this direction to calculating the desired signal function $\hat{s}_p(t)$ as $s_p(t) = \beta_p s_d(t - t_p)$. From the pseudo-spectrum of MUSIC-like search in Fig.3(a), we can see three peaks corresponding to the time delay $\hat{\boldsymbol{\tau}} = [-3.2,\ 0,\ 3.5]ns$, which is the relative delays to the signal $\hat{s}_p(t)$ from $\hat{\theta}_p = -29.82°$. The spectrum peaks of the time delays should be positioned in $\boldsymbol{\tau} = [\tau_1 - \tau_p, \tau_2 - \tau_p, \tau_3 - \tau_p]$. Since the minimum value in $\hat{\boldsymbol{\tau}}$ is $\tau_1 - \tau_p = -3.2ns$, we can know that real time delays of the signals are $[0,\ 3.2,\ 6.7]ns$ and the reference signal from $\hat{\theta} = -29.82°$ with time delay $3.2ns$ is not the direct signal. Consider the incident signal

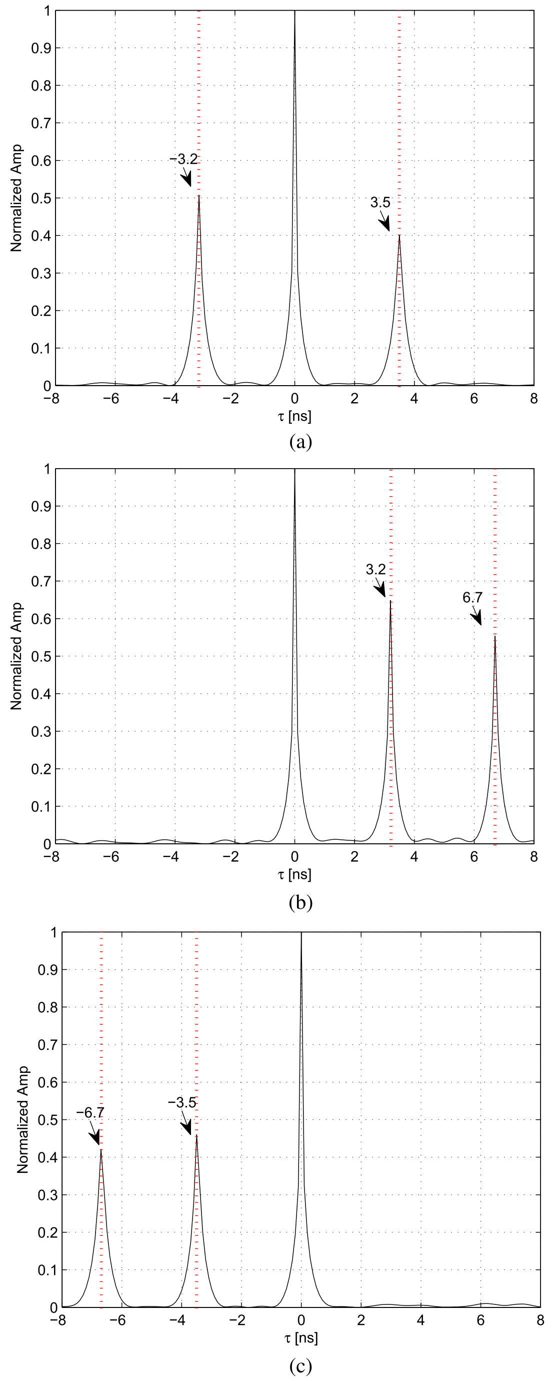

FIGURE 3. Pseudo-spectral of time delay estimation in SNR=20dB. (a) Signal from $\hat{\theta}_1 = -29.82°$ as the reference signal function. (b) Signal from $\hat{\theta}_1 = 14.97°$ as the reference signal. (c) Signal from $\hat{\theta}_1 = 60.39°$ as the reference signal.

from $\hat{\theta}_p = 14.97°$ as the reference signal, the relative delays to it are $\hat{\boldsymbol{\tau}} = [0,\ 3.2,\ 6.7]ns$ in Fig.3(b). The signal which comes from $\hat{\theta} = 14.97°$ is the direct signal, since the relative time delay in the three peaks are all positive in this spectrum. From the calculation above, we can also know that the signal of $\theta = 60.39°$ has the time delay $\tau = 6.7ns$. Thus, the TDOA and DOA of each signal is paired. Fig.3(c) shows that delay time of the signals from three directions relative to the third incident angle of $\theta = 60.34°$, which is not necessary for pairing the DOA and TDOA in general. In the Fig.4, we show Root Mean Square Error (RMSE) of DOA and time delay estimation for the signal with the Cramer-Rao bound (CRB) [6] versus SNR by 500 Monte Carlo trials.

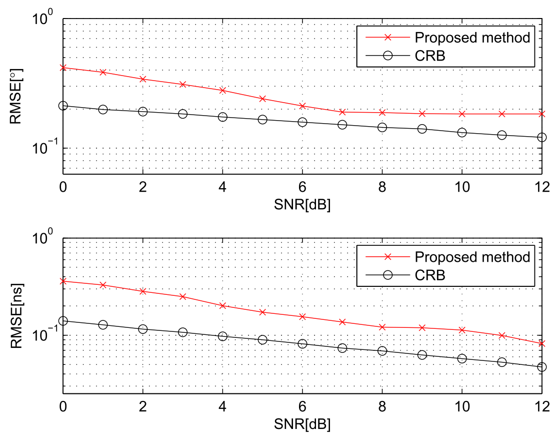

FIGURE 4. RMSE of DOA and TDOA for the three signal versus SNR.

(b) In order to further illustrate the DOA estimation performance of the proposed algorithm, it is compared with MUSIC [17], spatial smoothing MUSIC [20] and Topelize algorithm [21] in the separation ability for coherent signals. It is also assumed that the number of sources is an prior knowledge and the ULA of 10-element is divided into 7 overlapped sub-arrays (each sub-array have 4 elements) for smoothing MUSIC. In order to show clearly the performance of these algorithms, we reset the incident DOAs of the multipath signals. In Fig.5(a), the incident angle is $\theta = [10°, 13°, 60°]$ with the time delay $[0, 3.2, 6.7]ns$. We can see that MUSIC algorithm can't separate the coherent signals at all; Toeplize algorithm can barely distinguish between two adjacent signals. By comparison, the spatial smoothing and proposed algorithm can clearly distinguish all the signals. With the angle of incidence between the two coherent signals closing, Toeplize algorithm failed in Fig.5(b) and smoothing MUSIC in Fig.5(c). It can be seen that the proposed algorithm can form a substantially deeper null between the two closely spaced sources. Fig.6 shows the performance of the probability of separation with regard to the incident angle difference of two sources in SNR=20dB for the four methods . As we expected, the MUSIC algorithm for separating adjacent spatial sources is invalid when the sources with a strong temporal correlation, other algorithms are able to separate the adjacent sources. And proposed method exhibits the remarkable DOA separation ability among the methods.

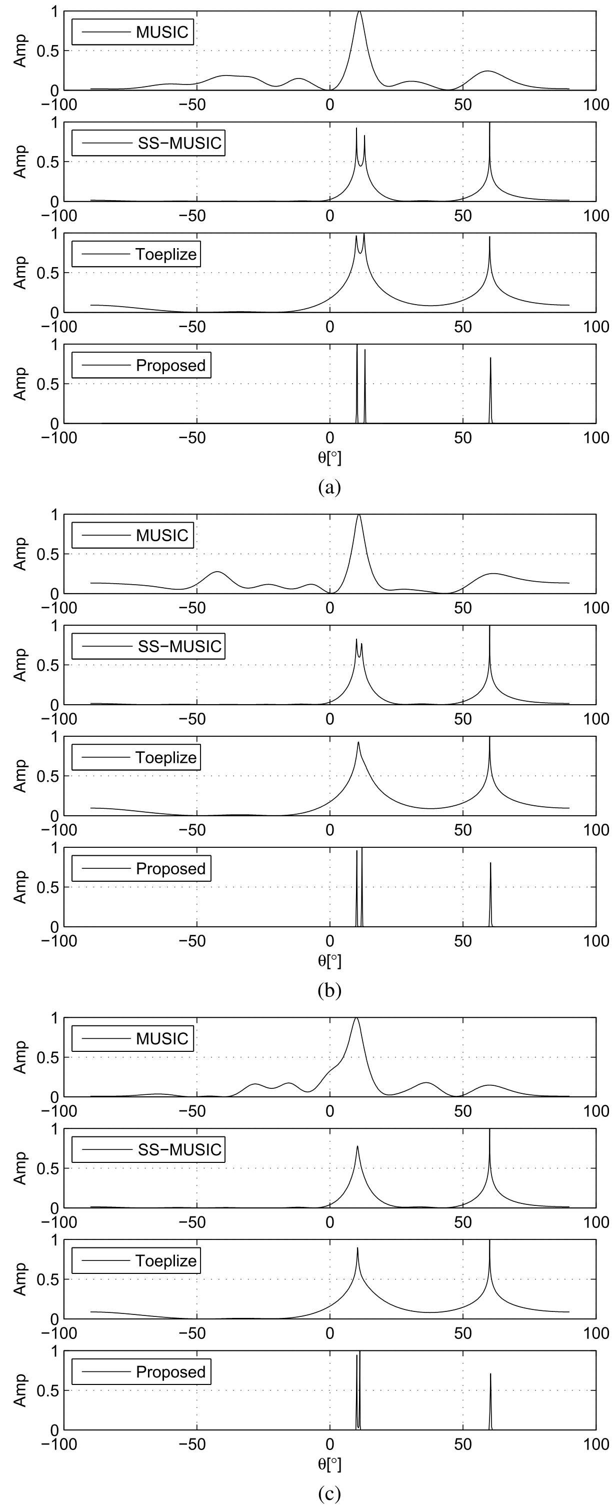

FIGURE 5. DOA separation for correlated sources in SNR=20dB. (a) $\theta = 10°, 13°, 60°$ with $\tau = 0ns, 3.2ns, 6.7ns$. (b) $\theta = 10°, 12°, 60°$ with $\tau = 0ns, 3.2ns, 6.7ns$. (c) $\theta = 10°, 11°, 60°$ with $\tau = 0ns, 3.2ns, 6.7ns$.

(c) We compare RMSE of the delay estimates of the proposed method, the Minimum Variance (MV) [22] and the

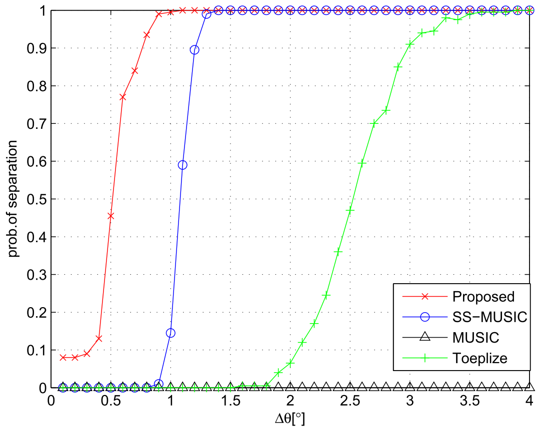

FIGURE 6. Probability of separation for correlated signals versus $\Delta\theta$.

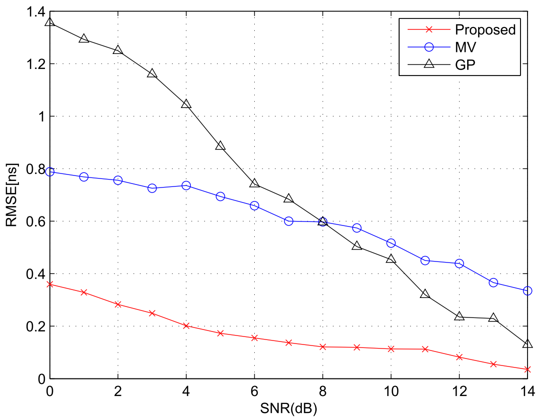

FIGURE 7. RMSE of time delay versus SNR.

GP [11] (which utilizes the geometrical properties of the signal in the frequency domain alternately for the time delay variables) algorithms. We reset the incident angles and the corresponding time delays of multipaths same to (a). In Fig.7, it shows RMSE of time delay estimation of the signal from $\theta = 60^\circ$ as an example with respect to the SNR. The RMSE value is calculated for 500 Monte Carlo simulations. The RMSE of the time delay estimated by the proposed algorithm is less than by utilizing the MV algorithm or GP algorithm. It is proved that the estimation of the proposed method is more effective.

VI. CONCLUSION

The paper proposes a novel algorithm which combines RISR technique along with the LCMP beamforming techniques and MUSIC-like searching algorithm to jointly estimate the DOAs and TDOAs under the multipath propagation conditions. The algorithm transforms the multi-dimensional estimation involved into two simple one-dimensional estimation. The DOA estimation process has super high-resolution based on an iterative scheme, which can be used to separate the relevant signals to any known array manifold without spatial smoothing. According to the obtained RISR spectrum, the direction of each relevant signal source can be distinguished and the number of the signal sources can be determined clearly. MUSIC-like searching process is used to obtain time delays of the sources by the phase shifts in frequency domain, so time delay within a fraction of the sampling time can be estimated. Monte Carlo simulations show the proposed algorithm has good performance for joint DOA and TOA estimation.

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