SCIENCE CHINA
Information Sciences

• LETTER •

May 2022, Vol. 65 159302:1–159302:2
https://doi.org/10.1007/s11432-021-3314-4

Fast ambiguous DOA elimination method of DOA measurement for hybrid massive MIMO receiver

Baihua SHI[1], Xinyi JIANG[1], Nuo CHEN[1], Yin TENG[1], Jinhui LU[1], Feng SHU[2*], Jun ZOU[1], Jun LI[1] & Jiangzhou WANG[3]

[1]*School of Electronic and Optical Engineering, Nanjing University of Science and Technology, Nanjing* 210094*, China;*
[2]*School of Information and Communication Engineering, Hainan University, Haikou* 570228*, China.;*
[3]*School of Engineering and Digital Arts, University of Kent, Canterbury* CT2 7NT*, UK*

Received 8 May 2021/Revised 27 July 2021/Accepted 9 August 2021/Published online 23 March 2022

Citation Shi B H, Jiang X Y, Chen N, et al. Fast ambiguous DOA elimination method of DOA measurement for hybrid massive MIMO receiver. Sci China Inf Sci, 2022, 65(5): 159302, https://doi.org/10.1007/s11432-021-3314-4

Dear editor,

Direction of arrival (DOA) estimation has been widely used in many applications, including wireless communications, radar, navigation, sonar, tracking of various objects, secure and precise wireless transmission (SPWT), rescue and other emergency assistance equipment [1,2]. In recent years, DOA estimation for massive multiple-input multiple-output (MIMO) system has attracted a lot of attention, which can provide ultra-high-resolution angle estimation. However, as the number of antennas tends to large-scale, due to its high computational complexity and circuit cost, it is difficult for massive MIMO to be widely used in DOA measurement. To address this issue, in [3], some low-complexity methods for the hybrid analog digital (HAD) structure were proposed, and the corresponding Cramer-Rao lower bound (CRLB) was derived. A novel discrete fourier transform (DFT)-based method was proposed in [4] to achieve the CRLB. As the artificial intelligence emerged, many machine learning methods have been integrated into the wireless communications [5]. A novel framework of combining deep-learning and massive MIMO was proposed in [6] to realize super-resolution channel estimation and DOA estimation. A low-complexity deep-learning-based method for a HAD MIMO system was proposed in [7], which can achieve similar or even lower normalized mean square error with much less complexity than the maximum likelihood method. Afterwards, authors in [8] proposed a new estimation of signal parameters via rotational invariance technique (ESPRIT)-based method and machine learning framework for the HAD structure. For a HAD MIMO system, there are two stages in the DOA measurement: DOA estimation of generating a set of candidate solutions and cancelling spurious solutions. Thus, the major challenging problem is how to eliminate direction-finding ambiguity rapidly. A smart strategy of maximizing the average receive power was proposed to remove $M-1$ spurious solutions in [3], where M is the number of antennas per subarray. This means it requires about $M-1$ time slots to infer the true direction angle with each time slot being multiple snapshots or samples. This means a large processing delay of M time slots. In this study, a fast ambiguous phase elimination method is proposed to find the true direction using only two-data-blocks with a slight performance loss. Our method achieves lower computational complexity and less estimated time at the cost of a little performance loss.

System model. The HAD antenna array captures the narrowband signal $s(t)\mathrm{e}^{\mathrm{j}2\pi f_c t}$ from the θ_0 direction emitted by a far-field transmitter, where $s(t)$ is the baseband signal and f_c is the carrier frequency. Here, a uniformly-spaced linear array (ULA) with N antennas is divided into K subarrays with each subarray containing M antennas, where $N = MK$. Via analog beamforming (AB), radio frequency (RF) chains, analog-to-digital convertors (ADCs), and digital beamforming (DB), the receive signal is $r^b(n) = \boldsymbol{v}_D^{\mathrm{H}}\boldsymbol{V}_A^{\mathrm{H}}\boldsymbol{a}(\theta_0)s(n) + \boldsymbol{v}_D^{\mathrm{H}}\boldsymbol{V}_A^{\mathrm{H}}\boldsymbol{w}^b(n)$, where b denotes the index of time slots, each time slot consists of L snapshots, $\boldsymbol{w}^b(n) \sim \mathcal{CN}(0, \sigma_w^2\boldsymbol{I}_M)$ is an additive white Gaussian noise (AWGN), $\boldsymbol{a}(\theta_0)$ is an array manifold, the DB vector is $\boldsymbol{v}_D = [v_1, v_2, \ldots, v_K]^{\mathrm{T}}$, and the AB matrix $\boldsymbol{V}_A$ is a block diagonal matrix. Let us define $\varphi = \frac{2\pi}{\lambda}d\sin\theta_0$, where λ and d denote the signal wavelength and antenna spacing, respectively.

Conventional Root-MUSIC-HDAPA algorithm. In the first stage, when all AB phases are zero, the output vector of sample n in time slot b is $\boldsymbol{y}_{\mathrm{AB}}^b(n) = M^{-\frac{1}{2}}\boldsymbol{a}_D(\theta_0)s^b(n) + \boldsymbol{w}_{\mathrm{AB}}^b(n)$, where $\boldsymbol{a}_D(\theta_0) = g(\theta_0)\boldsymbol{a}_M(\theta_0)$, $\boldsymbol{a}_M(\theta_0) = [1, \mathrm{e}^{\mathrm{j}M\varphi}, \ldots, \mathrm{e}^{\mathrm{j}(K-1)M\varphi}]^{\mathrm{T}}$, $g(\theta_0) = \sum_{m=1}^{M}\mathrm{e}^{\mathrm{j}(m-1)\varphi}$. The set of candidate solutions to DOA is estimated by using the Root-MUSIC algorithm. The sample covariance matrix of the output vector of the antenna array is $\boldsymbol{R}_{yy}^b = 1/L\sum_{n=1}^{L}\boldsymbol{y}_{\mathrm{AB}}^b(n)\boldsymbol{y}_{\mathrm{AB}}^b(n)$, whose singular-value decomposition (SVD) is expressed as $\boldsymbol{R}_{yy} = [\boldsymbol{E}_S\ \boldsymbol{E}_N]\sum[\boldsymbol{E}_S\ \boldsymbol{E}_N]^{\mathrm{H}}$ where $\boldsymbol{E}_S$ and $\boldsymbol{E}_N$ correspond to signal and noise subspaces, respectively. So the corresponding spectral function is $P_{\mathrm{MU}}(\theta) = \|\boldsymbol{a}_D^{\mathrm{H}}(\theta)\boldsymbol{E}_N\boldsymbol{E}_N^{\mathrm{H}}\boldsymbol{a}_D(\theta)\|^{-1}$. Let us define the poly-

* Corresponding author (email: shufeng0101@163.com)

nomial equation: $f_\theta(\theta) = \boldsymbol{a}_D^{\rm H}(\theta)\boldsymbol{E}_N\boldsymbol{E}_N^{\rm H}\boldsymbol{a}_D(\theta) \triangleq f_z(z) \triangleq f_\phi(\phi) = 0$, where $z = {\rm e}^{{\rm j}M\varphi}$ and $\phi = M\varphi$. The polynomial equation $f_z(z)$ has $2K-2$ roots z_i, which yields a set of associated emitter phases $\hat{\Theta}_r = \{\hat{\phi}_{r,i}, i \in \{1,2,\ldots,2K-2\}\}$. Digital phase alignment (DPA) is used to delete $2K-3$ pseudo solutions in $\hat{\Theta}_r$ and $\hat{\phi}_r$ is obtained. Then we can get $\hat{\phi}_r = 2\pi\lambda^{-1}Md\sin\hat{\theta}_r$. Since the function $f_\phi(\phi)$ is a periodic function of ϕ with period 2π, therefore, the extended feasible solution set is $\hat{\Theta} = \{\hat{\theta}_i, i \in \{0,1,\ldots,M-1\}\}$, where $\hat{\theta}_i = \arcsin(\frac{\lambda(\hat{\phi}_r+2\pi i)}{2\pi Md})$. Finally, analog phase alignment (APA) is used to eliminate the spurious solutions in the feasible set $\hat{\Theta}$.

Considering that the analog signal cannot be stored before ADCs, the new $M-1$ time slots should be received to eliminate $M-1$ spurious direction ambiguity in [3]. This will lead to a large estimation delay. To address this problem, a fast ambiguous phase elimination method is proposed to eliminate the spurious solutions by using only single time slot.

Proposed fast method of removing spurious solutions. Figure 1(a) shows the basic idea of eliminating spurious directions with $K \geqslant M$ in the second time slot. The total number K of subarrays are categorized into M groups of subarrays where each group has $P = K/M$ subarrays. In this slot, the phases of receive APA are designed according to M ambiguous directions such that all phases of the subarray group corresponding to the true direction are aligned to output the maximum power after APA, the output signal of the pth subarray of group m is as follows:

$$y_{mp}(n) = \boldsymbol{v}_{A,mp}^{\rm H}\boldsymbol{a}_{mp}(\theta_0)s(n) + w_{mp}(n), \tag{1}$$

where

$$\boldsymbol{v}_{A,mp} = \frac{1}{\sqrt{M}}[{\rm e}^{{\rm j}\alpha_{mp,0}}, {\rm e}^{{\rm j}\alpha_{mp,1}}, \ldots, {\rm e}^{{\rm j}\alpha_{mp,M-1}}], \tag{2}$$

where

$$\alpha_{mp,i} = \frac{2\pi}{\lambda}(H+i)d\sin\hat{\theta}_m, \tag{3}$$

where $H = (m-1)PM + (P-1)M$. The DB vector is set to be $\boldsymbol{v}_D = [1,1,\ldots,1]^{\rm T}$. Therefore, the output signal through DPA is $r_m(n) = \sum_{p=1}^{P} y_{mi}(n)$, and the average output power is

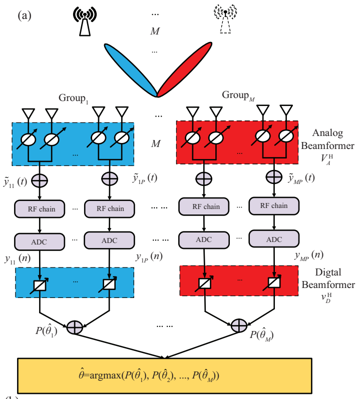
$$P_r(\hat{\theta}_m) = \frac{1}{L}\sum_{n=1}^{L}[r_m(n)r_m(n)^{\rm H}] = \frac{1}{L}\boldsymbol{r}\boldsymbol{r}^{\rm H}, \tag{4}$$

where $\boldsymbol{r} = [r(1),\ldots,r(L)]$. Eventually, the true direction angle corresponding to the maximum average power is

$$\hat{\theta} = \arg\max_{\hat{\theta}_m \in \hat{\Theta}} P_r(\hat{\theta}_m), \tag{5}$$

which completes the cancellation of spurious angles of requiring only one time slot. The delay is significantly reduced compared to the existing root-MUSIC-HDAPA DOA estimator method in [3]. The ratio of their total time delays is $2/M$. As M increases, the rapid advantage of the proposed method over root-MUSIC-HDAPA is more dramatic.

Computational complexity analysis. The computational complexities of the existing method and proposed method are $C_{\rm original} = \mathcal{O}(K^2L + (2(K-1))^3 + L((2K-2)K+NM))$, $C_{\rm proposed} = \mathcal{O}(K^2L + (2(K-1))^3 + L((2K-2)K+N))$ float-point operations (FLOPs). We can know that the computational complexity of the proposed method is reduced by M times when the ambiguous phase is eliminated.



(b)

SNR (dB)	RMSE (proposed method)	RMSE (root-MUSIC-HDAPA)
−20	32.9	29.3
−15	29.9	25.3
−10	21.5	18.4
−5	7.2	5.7
0	0.19	0.18
5	0.081	0.079
10	0.043	0.043

Figure 1 (Color online) (a) Proposed fast structure of removing spurious direction angles; (b) RMSE vs. SNR for different methods.

Simulation. Simulation parameters are chosen as follows: the direction of emitter $\theta_0 = 41.345°$, $N = 64$, $M = 4$, $L = 8$.

Figure 1(b) illustrates the performances of root mean square error (RMSE) versus SNR of the proposed method and the existing method in [3]. It can be seen from Figure 1(b) that the proposed method has wore performance. However, both methods achieve the similar performance at high SNR.

Acknowledgements This work was supported by National Natural Science Foundation of China (Grant Nos. 62071234, 61771244, 61901121) and Scientific Research Fund Project of Hainan University (Grant Nos. KYQD(ZR)-21007, KYQD(ZR)-21008).

References

1 Tuncer T E, Friedlander B. Classical and Modern Direction-of-Arrival Estimation. Burlington: Academic, 2009
2 Shu F, Wu X M, Hu J S, et al. Secure and precise wireless transmission for random-subcarrier-selection-based directional modulation transmit antenna array. IEEE J Sel Areas Commun, 2018, 36: 890–904
3 Shu F, Qin Y L, Liu T T, et al. Low-complexity and high-resolution DOA estimation for hybrid analog and digital massive MIMO receive array. IEEE Trans Commun, 2018, 66: 2487–2501
4 Fan D, Gao F F, Liu Y W, et al. Angle domain channel estimation in hybrid millimeter wave massive MIMO systems. IEEE Trans Wirel Commun, 2018, 17: 8165–8179
5 Yu X X, Guo J J, Li X, et al. Deep learning based user scheduling for massive MIMO downlink system. Sci China Inf Sci, 2021, 64: 182304
6 Huang H J, Yang J, Huang H, et al. Deep learning for super-resolution channel estimation and DOA estimation based massive MIMO system. IEEE Trans Veh Technol, 2018, 67: 8549–8560
7 Hu D, Zhang Y H, He L H, et al. Low-complexity deep-learning-based DOA estimation for hybrid massive MIMO systems with uniform circular arrays. IEEE Wirel Commun Lett, 2020, 9: 83–86
8 Zhuang Z H, Xu L, Li J Y, et al. Machine-learning-based high-resolution DOA measurement and robust directional modulation for hybrid analog-digital massive MIMO transceiver. Sci China Inf Sci, 2020, 63: 180302