AoA Estimation in Dual-Polarization Virtual Sub-Arrays for Analog Beam Refinement

Michel G. dos Santos, Laszlon R. Costa, Igor M. Guerreiro, Fredrik Athley

***Abstract*—This work proposes an angle of arrival (AoA) estimator for analog beam refinement in cellular systems operating in millimeter wave (mmWave) bands. Differently from previous works, the proposed method exploits the polarization domain and the concept of virtual arrays to estimate the AoA through a cross-correlation function. The proposed AoA estimator requires only a polarization-multiplexed measurement to estimate the AoA at the receiver. To cope with polarization leakage issues, this work proposes a channel equalization in the polarization domain conditioned on the estimation of the cross-channel gain. Simulation results indicate that the beam refinement procedure with the proposed AoA estimator has similar performance to the classical beam sweeping, but using less time-multiplexed measurements.**

***Index Terms*—Beam management, mmWave bands, Dual-polarization beamforming, AoA estimation.**

I. INTRODUCTION

Obtaining terabit per second data rates is a key requirement for sixth generation (6G), motivating the use of millimeter wave (mmWave) [1] bands. Large antenna arrays are then needed to avoid inherent severe propagation losses in such bands [2], which could limit network coverage.

In this context, beam management (BM) procedures ensure good beam pair links (BPLs). Initial BPLs usually comprise wide beams and are established during initial/random access phases. Then, once the user equipment (UE) is connected, its initial BPL can be boosted. The classical beam sweeping procedure (BSP) is an effective BM choice by sounding multiple BPL combinations. However, it usually consumes too many time-multiplexed reference signals (RSs) measurements [3].

The knowledge of dominant path angles enables faster beam refinements, reducing RS overhead. Accordingly, angle of arrival (AoA) estimators have been proposed [4]–[7]. In [4], the array power pattern is approximated by a Gaussian function and BSP is applied to estimate the AoA. In [5], the AoA estimator uses multiple single-polarized physical sub-arrays each connected to a single radio frequency (RF) chain, exploiting the phase correlation between the signal received by such subarrays. Similarly, [6] adopts an interleaved subarray architecture at the receiver and forms a monopulse signal from subarray responses to derive an AoA estimator. Finally, the AoA estimator [7] leverages phase correlation between virtual subarrays, avoiding the need for multiple physical subarrays.

Some limitations of such methods must be pointed out. In [4], the proposed method assumes an ideal array to approximate the array response function to a Gaussian function, ignoring the side lobe effects and realistic antenna gain patters. Although the methods proposed in [5], [6] require only a single transmission, it explores space diversity at the expense of requiring at least two physical sub-arrays with one RF chain each, which may imply increased hardware complexity. In [7], the authors choose to explore time diversity at the cost of requiring two time-multiplexed RS transmissions.

Given the limitations addressed above, the contributions of this paper may be summarized as follows:

- Dual-polarized virtual sub-arrays are used to obtain AoA estimates via phase correlation.
- To cope with polarization leakage, this work also proposes a cross-channel equalization procedure to enable the AoA estimation.

Michel G. dos Santos (orcid: 0009-0006-8163-2115), Laszlon R. Costa (orcid: 0000-0002-6700-0545) and Igor M. Guerreiro (orcid: 0000-0002-2157-3022) are with the Wireless Telecommunications Research Group (GTEL), Federal University of Ceará, Fortaleza, Brazil (e-mails:michel.gonzaga, laszlon, igor@gtel.ufc.br). Laszlon R. Costa is also with the Federal University of Cariri (UFCA), Juazeiro do Norte, Brazil. Fredrik Athley (orcid: 0009-0002-2385-4140) is with Ericsson Research, Ericsson, Sweden (e-mail: fredrik.athley@ericsson.com)

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- To deal with phase ambiguity in the AoA estimate, this work proposes a sub-optimal solution given some previous beam direction.

Simulation results show the proposed method achieves comparable angle estimation accuracy to [7] using less time-multiplexed measurements, and is robust against estimation errors.

Notations: $(\cdot)^H$, $(\cdot)^T$, $(\cdot)^*$ denote Hermitian, transpose, and conjugate, respectively. $\mathrm{E}[\cdot]$ is expectation. Bold letters $\mathbf{x}$ and $\mathbf{X}$ represent vectors and matrices. $\mathcal{W}_x$ is a codebook of x-dimensional unit-norm beams. Angles $(\cdot)'$ are defined locally.

II. System Model

Consider a transmitter (Tx)/receiver (Rx) pair in a multiple-input multiple-output (MIMO)-orthogonal frequency division multiplexing (OFDM) cellular network operating at carrier frequency f_c in a higher-frequency band[1]. The development described in the following applies to uplink or downlink.

Both Tx and Rx use dual-polarized M_T and M_R element uniform linear arrays (ULAs), respectively, vertically placed, with linear slant angles $\xi_{T,A}, \xi_{T,B}$ and $\xi_{R,A}, \xi_{R,B}$, respectively. Analog beamforming is deployed across the antenna elements (AEs) connected to each digital port, controlling only the phase of their excitation weights and keeping their amplitude constant. AE spacing equals $d = 0.5\lambda$. The array response vectors are given by:

$$\mathbf{a}_R(\theta) = \left[e^{j(M_R-1)\psi_{R,z}} \quad \ldots e^{jn\psi_{R,z}} \quad \ldots \quad 1\right]^T, \quad (1)$$

$$\mathbf{a}_T(\theta) = \left[e^{j(M_T-1)\psi_{T,z}} \quad \ldots e^{jm\psi_{T,z}} \quad \ldots \quad 1\right]^T, \quad (2)$$

with $\psi_{R,z} = \psi_{T,z} = 2\pi d \cos\theta$.

Assuming a single-path channel, typical at higher frequency bands due to channel sparsity [8], let θ_{AoD} and ϕ_{AoD} denote the angles of departure from the Tx ULA. The MIMO channel matrix $\mathbf{H}_{i,j} \in \mathbb{C}^{M_R \times M_T}$ between digital ports $i, j \in \{A, B\}$ is then given by

$$\mathbf{H}_{i,j} = \gamma_{i,j}\mathbf{a}_R(\theta_{AoA})\mathbf{a}_T(\theta_{AoD})^T, \quad (3)$$

where $\gamma_{i,j} = \sqrt{A_{\mathrm{RPP}}}\frac{\lambda}{4\pi d}\cos(\xi_{R,i} + \xi_{T,j})$, $A_{\mathrm{RPP}} \triangleq A\left(\theta_{AoA}, \phi_{AoA}\right) A\left(\theta_{AoD}, \phi_{AoD} - \alpha\right)$, where $A(\theta, \phi)$ is the antenna radiation pattern defined in [9], which is a function of the azimuth and elevation angle. Angle α is the rotation around the z-axis. For simplicity, the channel in (3) is assumed constant

[1]Single-carrier might be assumed instead under the cost of eventual performance losses due to fewer available samples.

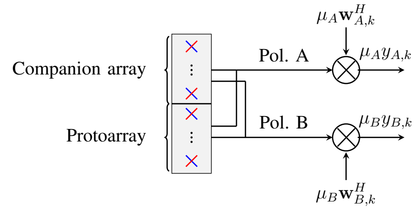


Figure 1. Virtual sub-array representation: protoarray (bottom) and companion array (top).

within a few consecutive time resources. The received signal $y_{i,k} \in \mathbb{C}$ at digital port i and receive beam k is:

$$y_{i,k} = \sqrt{\frac{P_{tx}}{2}} \sum_{j\in\{A,B\}} \mathbf{w}_{i,k}^H \mathbf{H}_{i,j}\mathbf{f}_j^* s + z_{i,k}, \quad (4)$$

where s is the transmitted symbol, $\mathbf{w}_{i,k}, \mathbf{f}_j$ are beamforming vectors, and $z_{i,k}$ is AWGN. From (3), this simplifies to:

$$y_{i,k} = \mathbf{w}_{i,k}^H \mathbf{a}_R(\theta_{AoA}) b_i + z_{i,k}, \quad (5)$$

with b_i grouping transmit-side parameters.

Virtual sub-arrays: Assuming even M_R, the receive array is divided into two halves. The array response then becomes:

$$\mathbf{a}_R(\theta) = \left[e^{j\frac{M_R}{2}\psi_{R,z}}\mathbf{a}_0(\theta)^T \quad \mathbf{a}_0(\theta)^T\right]^T. \quad (6)$$

Each half-size array is referred to as a *virtual sub-array*. As depicted in Figure 1, we denote the top and bottom sub-arrays as companion array and protoarray, respectively. Consequently, $\mathbf{a}_0\left(\theta\right)$ stands for the array response of the protoarray.

A. Problem Formulation

Given an initial BPL obtained during some coarse BM procedure, the goal is to perform a beam refinement on the receiver side, i.e., to select $\mathbf{w}_{A,k}, \mathbf{w}_{B,k} \in \mathcal{W}_{M_R}$ in Eq. (4) that boost the BPL quality in terms of received power. To this end, the transmit beams $\mathbf{f}_A, \mathbf{f}_B \in \mathcal{W}_{M_T}$ in (4) are assumed fixed.

From (5), we may define the equivalent channel $\Delta_i(\mathbf{w}_{i,k})$ at the i-th port and beamformed by $\mathbf{w}_{i,k}$ as $\Delta_i(\mathbf{w}_{i,k}) = \sum_{j\in\{A,B\}} \mathbf{w}_{i,k}^H \mathbf{H}_{i,j}\mathbf{f}_j^*$, which can be estimated by assuming that symbol s is known and transmitted in multiple subcarriers where channel $\mathbf{H}_{i,j}$ is constant for the purpose of noise mitigation. The BPL quality is then evaluated via the bijective function $\Delta(\mathbf{w}_{A,k}, \mathbf{w}_{B,k}) = \sum_{i\in\{A,B\}} |\Delta_i(\mathbf{w}_{i,k})|^2$.

Finally, let $\mathcal{W} \subset \mathcal{W}_{M_R}$ be a codebook of M_R-dimensional candidate beam vectors that contains $L \leq |\mathcal{W}_{M_R}|$ vectors of $\mathcal{W}_{M_R}$. The beam refinement problem can then be formulated as

$$\max_{\mathbf{w}_A, \mathbf{w}_B \in \mathcal{W}} \Delta(\mathbf{w}_A, \mathbf{w}_B)\,. \tag{7}$$

The problem in (7) is combinatorial as $\mathcal{W}$ is finite and its solution can be found via exhaustive search. Note that the smaller L the lower the exhaustive search burden, but less beam directions are sounded, which decreases the chance of boosting the BPL.

From (5), the BPL quality can be redefined in terms of the gain $|\mathbf{w}^H \mathbf{a}_R(\theta_{AoA})|^2$, assuming that the same beamformer vector is used in both digital ports, i.e., $\mathbf{w}_A = \mathbf{w}_B = \mathbf{w} \in \mathcal{W}_{M_R}$. Although θ_{AoA} cannot be known directly, its estimate $\hat{\theta}_{AoA}$ can be obtained. A problem similar to (7) can be formulated as $\max_{\mathbf{w} \in \mathcal{W}_{M_R}} \left| \mathbf{w}^H \mathbf{a}_R(\hat{\theta}_{AoA}) \right|^2$,which scales with the cardinality of $\mathcal{W}_{M_R}$.

III. Proposed angle estimator, DP-ViSA

The proposed AoA estimation framework extends the virtual-subarray based AoA (ViSA) algorithm [7], both exploiting phase shifts between sub-arrays. Unlike ViSA, which cross-correlates time-lagged RS, our method uses polarization-multiplexed measurements with cross-channel equalization in the polarization domain.

Proposition 1. *Let y_i be the received signals at each polarization, $i \in \{A, B\}$. The phase shift $e^{j\frac{M_R}{2}\psi_{R,z}}$ is acquired exploiting the cross-correlation between*

$$r_d = \mu_B y_B - \mu_A y_A, \tag{8}$$
$$r_s = \mu_B y_B + \mu_A y_A, \tag{9}$$

where $\mu_A = 1$, $\mu_B = b_A/b_B$, and, as in [10],

$$\mathbf{w}_A = \frac{1}{\sqrt{2}} \begin{bmatrix} -\mathbf{w}_0^T & \mathbf{w}_0^T \end{bmatrix}^T, \mathbf{w}_B = \frac{1}{\sqrt{2}} \begin{bmatrix} \mathbf{w}_0^T & \mathbf{w}_0^T \end{bmatrix}^T, \tag{10}$$

with $\mathbf{w}_0 \in \mathcal{W}_{M_R/2}$, and used to estimate the AoA.

Proof. Using (6) and (10) in (5) yields:

$$y_A = \frac{1}{\sqrt{2}}(1 - e^{j\frac{M_R}{2}\psi_{R,z}})\mathbf{w}_0^H \mathbf{a}_0 b_A + z_A, \tag{11}$$
$$y_B = \frac{1}{\sqrt{2}}(1 + e^{j\frac{M_R}{2}\psi_{R,z}})\mathbf{w}_0^H \mathbf{a}_0 b_B + z_B. \tag{12}$$

Substituting into (8) and (9), we obtain:

$$r_d = \frac{2}{\sqrt{2}} e^{j\frac{M_R}{2}\psi_{R,z}} \mathbf{w}_0^H \mathbf{a}_0 b_A + z_B \frac{b_A}{b_B} - z_A, \tag{13}$$
$$r_s = \frac{2}{\sqrt{2}} \mathbf{w}_0^H \mathbf{a}_0 b_A + z_B \frac{b_A}{b_B} + z_A \tag{14}$$

Finally, the cross-correlation becomes:

$$\rho = (r_s)^* r_d = |\bar{b}|^2 e^{j\frac{M_R}{2}\psi_{R,z}} + g_z, \tag{15}$$

where $\bar{b} = \frac{2}{\sqrt{2}}\mathbf{w}_0^H \mathbf{a}_0 b_A$ and g_z is a noisy term. □

Phase estimation: From (15), a set of candidate phases is defined:

$$\mathcal{P}_{\psi_{R,z}} = \left\{ \frac{\angle\rho + 2k\pi}{M_R/2} \,\middle|\, k \in \{-M_R/2, M_R/2\} \right\}. \tag{16}$$

where $\mathcal{K} = \{-\frac{M_R}{2}, \frac{M_R}{2}\}$, then introducing a phase ambiguity problem, also observed in [7].

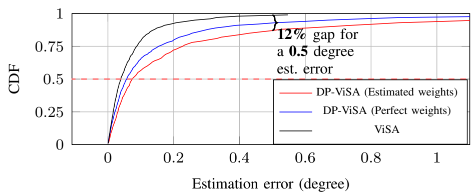
Let $\hat{\psi}_{R,z} \in \mathcal{P}_{\psi_{R,z}}$ be the estimated version of $\psi_{R,z}$. Since all the possible values of $\frac{2k\pi}{M_R/2}$ are known and given that $|\cos(\theta_{AoA})| \leq 1$, it is possible to obtain a subset of possible values of θ_{AoA}, defined as $\mathcal{P}_{\theta_{AoA}} = \left\{ \cos\left(\frac{\hat{\psi}_{R,z}}{2\pi d_z} - \frac{2k}{d_z M_R} \right)^{-1} \,\middle|\, \left| \frac{\hat{\psi}_{R,z}}{2\pi d_z} - \frac{2k}{d_z M_R} \right| \leq 1 \right\}$. Then, the best $\hat{\theta}_{AoA}$ may be selected by solving the following problem: $\hat{\theta}_{AoA} = \arg\max_{\theta \in \mathcal{P}_{\theta_{AoA}}} \left| (\mathbf{w}^\star)^H \mathbf{a}_R(\theta, \phi_{AoA}) \right|^2$, where $\mathbf{w}^\star$ is the wide beam from the initial BPL.

IV. Numerical Results

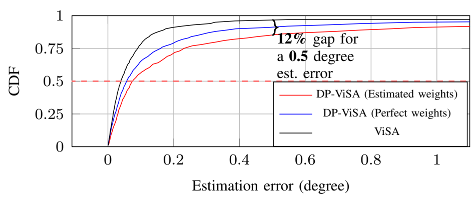
In this section, numerical results are provided to evaluate the proposed beam refinement procedure.

The Tx is randomly positioned using a uniform distribution within the Rx's azimuth range in $[-\pi/3, \pi/3]$, with a Rx-Tx distance $d \in [10, 50]$ m. The zenith angles of the line-of-sight (LOS) path are derived from the Rx and Tx coordinates.

A total of 1000 Monte Carlo simulations were run. The Rx is placed at 10 m height, while the Tx is at 1.5 m. The Rx employs a ULA with $M_R = 64$ elements and the Tx uses $M_T = 8$ elements. $\alpha = \pi$ radians. Initial beam selection at the Rx uses wide beams [10] to find the beam with the highest reference signal received power (RSRP), while the Tx uses DFT beam 0. The transmit power is $P_{tx} = 10$ dBm and the noise power spectral density is -174 dBm/Hz. A DFT-based codebook

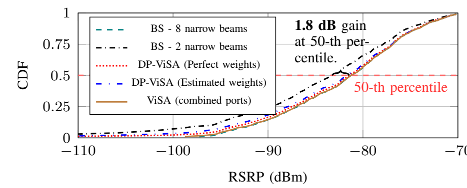


(a) Optimum ambiguity solution.

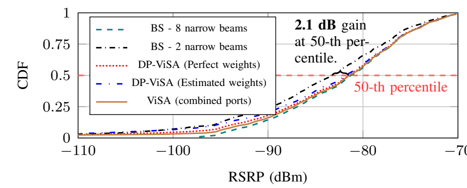


(b) Proposed ambiguity solution.

Figure 2. Absolute AoA estimation error's CDF.



(a) Optimum ambiguity solution.



(b) Proposed ambiguity solution.

Figure 3. RSRP's CDF.

without oversampling is used. The carrier frequency is $f_c = 100$ GHz and subcarrier spacing is 960 kHz. The channel model includes only the LOS component [9], and simulations use a comb-type structure with 12 frequency-multiplexed symbols[2] and a free-space path loss model.

The technique's quality was evaluated using two metrics: absolute AoA estimation error and BPL's RSRP. A simplified layer 1 (L1)-RSRP is calculated as $\sum_{i\in\{A,B\}} \left|\sum_{j\in\{A,B\}} \mathbf{w}_i^H \mathbf{H}_{i,j}\mathbf{f}_j\right|^2$.The proposal, herein called dual-polarization virtual subarray based AoA (DP-ViSA) was compared with (i) ViSA [7], and (ii) BSP using either two or eight narrow beams. The required number of measurements is: 1 for DP-ViSA, 2 for ViSA and two-beam BSP, and 8 for eight-beam BSP.

We compare DP-ViSA against ViSA, as both rely on similar transceiver hardware architectures based on virtual subarrays, herein assumed without any hardware imperfection for simplicity. ViSA differs by applying a Rx beamforming configuration similar to (10) over two time-multiplexed measurements.

To assess ambiguity suppression, DP-ViSA and ViSA are evaluated with both optimal and proposed methods, where $\hat{\psi}_{R,z}$ is compared to all candidate values in (16) and the closest is selected. Simulations also consider the impact of using either an estimated or exact value of the weight ratio $\frac{b_A}{b_B}$.

In Figs. 2a and 2b, the 50th-percentile performance of both ambiguity suppression methods is very similar. Although ViSA slightly outperforms DP-ViSA (estimated weights), the gap is merely 0.035 degrees, possibly negligible for beam selection. At a 0.5-degree error, the percentile difference reaches up to 12%, but DP-ViSA still maintains performance above the 80th percentile.

Figs. 3a and 3b show that DP-ViSA and ViSA yield similar RSRP distributions. Notably, DP-ViSA (estimated weights) outperforms 2 narrow beams BSP by nearly 2 dB at the 50th percentile. For both ambiguity solutions, DP-ViSA achieves performance comparable to eight-beam BSP, albeit slightly lower with the proposed ambiguity method, as expected due to its non-optimality.

V. Conclusion

This paper proposed a dual-polarization extension of the single-polarization ViSA [7] for analog beam refinement in cellular systems, which provides similar results, while requiring only a single measurement, conditioned on dual-polarization gain knowledge, obtained from previous signaling, e.g, during random access, to perform the AoA estimation.

Furthermore, the proposed method provided performance near an 8-long beam sweeping, but with only a single dual-polarization measurement, i.e., 8x less signaling. The proposal proved up to 2dB better than a 2-long beam sweeping. A sub-optimum solution, which delivers reliable results, for the ambiguity problem present in both ViSA and DP-ViSA, was also proposed as a contribution of this work.

As future work, one may assess DP-ViSA in rectangular arrays with different sub-array virtualization schemes, under hardware imperfection and non-LOS condition. Also, computational complexity may be investigated.

[2]Other comb-type factors can be employed w/o loss of generality.

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