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Motionless, Single-Snapshot Far-field Pattern and EIRP Measurement Method in a Compact Hybrid Chamber Employing a Sampling Array

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Abstract—A method for motionless far-field radiation pattern and Equivalent Isotropic Radiated Power (EIRP) measurement of black-box transmitters in a compact antenna chamber is presented. Motionless operation is particularly valuable for rapid production testing of complex devices under test (DUTs) such as beamsteering array antennas. The chamber is a metal overmoded rectangular waveguide terminated by a sampling chamber array antenna (CAA) at one end and an absorbing wall at the other end, where the DUT test zone is located. Using reciprocity and power-wave formalism, we derive a linear relation between the coherently sampled CAA receive signals and the DUT EIRP pattern, enabling pattern reconstruction from a single-snapshot without mechanical rotation. A one-time calibration procedure, based on the measured embedded element fields (EEFs) of the CAA and an optimal beamforming algorithm, yields CAA weight vectors that, in the transmit situation, synthesize a rich set of plane waves in the test zone, thereby defining a high-resolution sampling grid of the pattern and EIRP normalization. A dedicated probe-correction technique is developed to ensure accurate EEF estimation despite the strongly multi-scattering environment. Full-wave simulations in the 5G NR FR1 range and experiments in a 900 MHz prototype chamber validate the method for representative linear and circular polarized (LP/CP) DUT cases. The measured EIRP errors at peaks are 0.2–0.4 dB for the three experimental DUTs. The maximum pointwise normalized pattern errors within $\pm 60^\circ$ are between -17 dB and -25 dB, with the largest value associated with a localized feature in the anechoic chamber reference for the low-gain CP spiral DUT. In the high-gain simulated cases, the errors decrease below -30 dB.

Index Terms—Antenna measurement, radiation pattern, phased array, plane wave generator, beamforming

I. INTRODUCTION

OVER-THE-AIR (OTA) testing has become essential for characterizing modern antenna systems, enabling the evaluation of performance metrics such as antenna efficiency, total radiated power (TRP), equivalent isotropically radiated power (EIRP), receiver sensitivity, and far-field patterns under conditions that closely resemble practical deployments. Indoor

OTA testing is typically preferred over outdoor measurements because it provides a controlled environment free of external interference, improved repeatability, and reduced measurement uncertainty.

However, as modern systems evolve toward highly integrated, beam-steering and multi-antenna architectures and demand evaluation of a broad set of performance metrics, traditional OTA test methods face limitations. Anechoic chambers (ACs) and compact antenna test ranges (CATRs) [1] provide high-accuracy far-field pattern measurements but require large physical dimensions, mechanical rotation of the device under test (DUT), and long test cycles. Even in compact or tapered range configurations, the conditions under which absolute gain and EIRP can be reliably extracted remain an active topic of investigation and are known to depend on propagation assumptions and frequency-specific behavior [2]. Planar near-field scanning approaches [3], [4] can reduce chamber volume but require mechanical scanning over large surfaces and may restrict angular coverage. Reverberation chambers (RCs), on the other hand, offer fast and cost-efficient measurements of TRP and sensitivity metrics, but with lower accuracy for directional metrics without substantial post-processing and additional constraints [5]–[7].

Several hybrid or enhanced OTA test concepts have been proposed to mitigate these constraints. Plane-wave generators (PWGs) [8], [9] and multi-probe anechoic chambers (MPAC) [10]–[13] reduce test time and physical size but remain limited in angular coverage or test zone size. Recent efforts introduce multi-PWG architectures [1] or combine CATRs with RCs [14] to broaden coverage or create richer propagation environments for pattern extraction, yet they typically require costly hardware expansions, complex calibration procedures or provide limited accuracy.

To address these limitations, a hybrid (AC-RC) OTA chamber concept was proposed in [15], leveraging rich multi-mode propagation of an overmoded rectangular waveguide (WG) together with a chamber array antenna (CAA) to accurately synthesize plane wave fields at the DUT test zone (TZ). Prior studies [16]–[19] analyzed aspects of plane-wave synthesis and field controllability in this geometry, but a complete methodology for accurate far-field pattern or EIRP reconstruction has not been established. Furthermore, to the authors' knowledge, there is currently no compact OTA chamber that reconstructs the full complex far-field radiation pattern and absolute EIRP of black-box transmitters from a single coherent snapshot,

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without mechanical DUT rotation. This work establishes the theoretical foundation and experimental proof-of-concept of a fast, motionless far-field pattern and EIRP reconstruction method for black-box transmitters in a compact overmoded WG chamber, targeting measurement scenarios where compactness, no mechanical DUT rotation, and short per-DUT-state acquisition time are primary requirements. This method is enabled by three key theoretical developments:

- a reciprocity- and power-wave based formulation that derives, for the first time in this context, a linear mapping between the CAA coherently sampled output signals and the DUT EIRP pattern;
- a one-time chamber calibration procedure that uses measured embedded element fields and optimal CAA beamforming to synthesize a rich spectrum of plane waves in the TZ, thereby defining a high-resolution sampling grid of the pattern and EIRP normalization; and
- a modal probe-correction technique that reconstructs vector electric fields in the WG chamber by explicitly accounting for all propagating modes, enabling accurate estimation of embedded element fields in a highly multi-scattering WG environment.

These theoretical contributions are supported by comprehensive numerical validation for the reconstruction accuracy for a broad range of DUT patterns, including low- to high-gain cases and multiple scan angles. Experimental verification is performed using a 900 MHz prototype chamber with AC measurements used as the reference. The measurements were performed for three DUTs, whose physical sizes (and therefore their gains) are limited by the physical TZ size and the manufactured CAA prototype.

Sec. II presents the theoretical framework for the proposed method, including the mathematical formulation, the reciprocal measurement model, and the algorithmic procedure for reconstructing the DUT's EIRP pattern. Sec. II-D presents the chamber calibration procedure and Sec. III provides full-wave numerical validation of the EIRP pattern reconstruction. Sec. IV describes the probe-correction technique required for practical measurements, and Sec. V reports experimental results. Finally, conclusions are summarized in Sec. VII, and the derivation of the probe response to an incident plane wave is given in the Appendix.

II. RADIATION PATTERN MEASUREMENT METHODOLOGY

This section describes the proposed methodology for the far-field radiation pattern measurement of a transmitting DUT. Since we aim to characterize a “black-box” DUT without direct access to the antenna ports and having unknown total radiated power, the pattern will be normalized to represent the Equivalent Isotropic Radiated Power (EIRP).

A. Chamber configuration

The hybrid chamber configuration used for the pattern measurements is shown in Fig. 1. The WG dimensions are (W, H, L) . We found that, for the particular chamber dimensions in [16], the TZ size is $\approx 0.4W \times 0.4W$, which allows us to synthesize a wide angular range of plane waves (PWs)

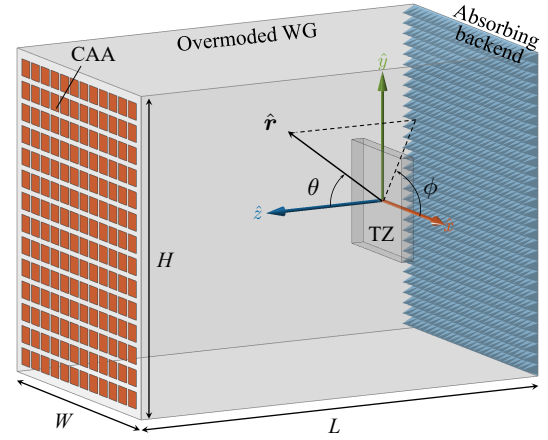


Fig. 1. The model of the hybrid chamber. It consists of a metal waveguide (WG) with dimensions (W, H, L) , and the CAA located at the left end of the WG. The test zone is denoted by the gray rectangle at the opposite end of the WG. The WG is terminated by the absorbing backend.

impinging on the TZ. The number of elements in the CAA, N_{el} , and their layout are determined by the highest frequency of interest for the pattern measurements, where the inter-element spacing should not exceed $0.7\lambda - 0.8\lambda$ while filling the entire WG cross-section. Since for the pattern measurements we will synthesize plane waves in the TZ, the array layout selection follows guidelines from [17].

B. Notations

By definition, the power wave exciting the DUT in the transmitting situation is [20], [21]

$$a^{\text{dut}} = \frac{V_t + Z_0 I_t}{2\sqrt{Z_0}} = \frac{(Z_{\text{in}}^{\text{dut}} + Z_0) I_t}{2\sqrt{Z_0}}, \quad (1)$$

where V_t and I_t are the terminal voltage and current, respectively (see Fig. 2); $Z_{\text{in}}^{\text{dut}}$ is the input impedance of the DUT antenna, and Z_0 is the reference impedance, which is chosen to be real-valued. Furthermore, let $P_{\text{av}} = \frac{1}{2}|a^{\text{dut}}|^2$ denote the available power from the RF backend of the DUT connected to the antenna port¹.

Consider Fig. 1, where the direction $\hat{\mathbf{r}} = [\sin \theta \cos \phi, \sin \theta \sin \phi, \cos \theta]$ is defined in the TZ coordinate system. The transmitting DUT's far-field function, when excited by a^{dut} , is $\mathbf{G}^t(\hat{\mathbf{r}}) = \mathbf{E}^t(\mathbf{r}) r e^{j2\pi r/\lambda}$, where r is the far-field distance at which the electric field $\mathbf{E}^t$ is defined. Let us define the vector realized gain pattern of the DUT by

$$\mathbf{G}_r(\hat{\mathbf{r}}) = \sqrt{\frac{2\pi}{\eta P_{\text{av}}}} \mathbf{G}^t(\hat{\mathbf{r}}) = G_{r,\text{co}}(\hat{\mathbf{r}}) \hat{\mathbf{c}}\mathbf{o} + G_{r,\text{xp}}(\hat{\mathbf{r}}) \hat{\mathbf{x}}\mathbf{p} \quad (2)$$

where $G_{r,\text{co}}(\hat{\mathbf{r}})$ and $G_{r,\text{xp}}(\hat{\mathbf{r}})$ are complex-valued partial realized gain patterns, normalized as in [22, Eq. (2.70)]. The polarization component will be denoted as $p \in \{\text{co}, \text{xp}\}$, where the two orthogonal components may represent either a linear polarization basis, such as the Ludwig-3 definition, or a circular polarization basis, such as LHCP/RHCP. In this

¹For fully-integrated devices, a reference plane somewhere between the RF backend and antenna can be considered as a “port”.

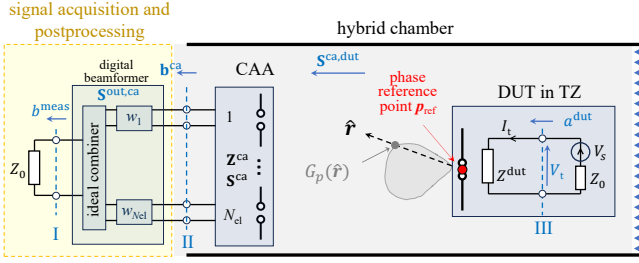


Fig. 2. Schematic of the chamber setup for the DUT pattern measurement.

way, $G_r^{\text{dut}}(\hat{r}) = |\mathbf{G}_r(\hat{r})|^2$ is the standard IEEE realized gain pattern of the DUT.

The above enables us to define the *EIRP pattern* as

$$\mathbf{G}(\hat{r}) = |a^{\text{dut}}| \mathbf{G}_r(\hat{r}) \quad (3)$$

where the excitation phase information is already included in $\mathbf{G}_r$. Upon substituting (2) in (3), one obtains the relation

$$\mathbf{G}^t(\hat{r}) = \frac{\mathbf{G}(\hat{r})}{2} \sqrt{\frac{\eta}{\pi}}. \quad (4)$$

The EIRP of a single-port² DUT is now defined as

$$\text{EIRP}(\hat{r}) = P_{\text{av}} G_r^{\text{dut}}(\hat{r}) = \frac{1}{2} |a^{\text{dut}} \mathbf{G}_r(\hat{r})|^2 = \frac{1}{2} |\mathbf{G}(\hat{r})|^2, \quad (5)$$

The measurement procedure for $\mathbf{G}(\hat{r})$ is derived in the following sub-section.

C. Method to measure far-field EIRP pattern

Consider the measurement system shown in Fig. 2 in which a transmitting DUT is placed in the TZ of the chamber. The DUT's antenna, excited internally at the reference plane III by the wave a^{dut} , radiates into the overmoded waveguide, and the radiated field is received by the CAA elements, digitized and combined by the beamformer to the single outgoing wave b^{meas} (reference plane I), which can be expressed as

$$b^{\text{meas}} = \mathbf{S}^{\text{out,ca}} \mathbf{S}^{\text{ca,dut}} a^{\text{dut}} = S^{\text{out,dut}} a^{\text{dut}}, \quad (6)$$

where $S^{\text{out,dut}} = \mathbf{S}^{\text{out,ca}} \mathbf{S}^{\text{ca,dut}}$ and $\mathbf{S}^{\text{out,ca}} = [w_1, \dots, w_{N_{\text{el}}}] / \sqrt{N_{\text{el}}}$ is the $1 \times N_{\text{el}}$ input-to-output coupling S-matrix of the beamformer; N_{el} is the number of antenna elements in the CAA; and $\mathbf{S}^{\text{ca,dut}}$ is the $N_{\text{el}} \times 1$ coupling S-matrix between the CAA and DUT port (between ref. planes II and III).

Now, let us consider a reciprocal case when the same DUT is receiving, see Fig. 3. In this case, the incident wave a excites the CAA through the beamformer with the same excitation coefficients $\mathbf{w}$ as the weight coefficients in Fig. 2. Then, the power wave received by the DUT is

$$b_r^{\text{dut}} = S^{\text{dut,out}} a. \quad (7)$$

Assuming reciprocity, $S^{\text{out,dut}} = S^{\text{dut,out}}$, and therefore, substituting (7) into (6), we get

$$b^{\text{meas}} = \frac{b_r^{\text{dut}}}{a} a^{\text{dut}}. \quad (8)$$

²In case of multi-antenna DUTs, a compound “beam” port (i.e. input of a beamformer) is considered as the single input of the antenna system.

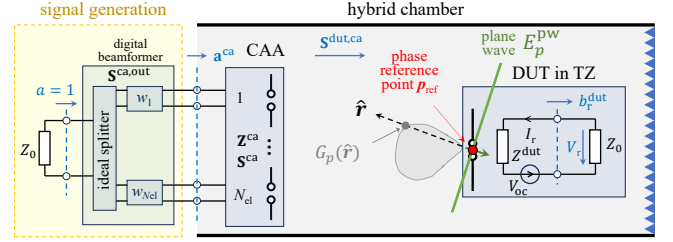


Fig. 3. Reciprocal case when DUT is receiving.

Since the goal is to measure a *far-field* of the DUT, it should “experience” a plane wave (PW) incoming from the direction of measurement. Therefore, the excitation coefficients $\mathbf{w}$ are chosen during the chamber calibration procedure (details follow in Sec. II-D) in such a way that the field radiated by the CAA forms a field distribution in the TZ corresponding to the PW impinging TZ from direction $\hat{r}$ and having E-field vector $\mathbf{E}^{\text{pw}}$ at the phase reference point $\mathbf{p}_{\text{ref}}$ (see Fig. 3). When the DUT is placed in the TZ, the induced open-circuit terminal voltage due to the incident PW is [22, Eq.(2.129)]

$$V_{\text{oc}} = -\frac{2j\lambda}{\eta I_t} \mathbf{G}^t(\hat{r}) \cdot \mathbf{E}^{\text{pw}}(-\hat{r}, \mathbf{p}_{\text{ref}}) \quad (9a)$$

$$= -\frac{2j\lambda}{\eta I_t} G_p^t(\hat{r}) E_p^{\text{pw}}(-\hat{r}, \mathbf{p}_{\text{ref}}), \quad (9b)$$

where η is the free-space impedance; λ is the wavelength; and $\mathbf{G}^t(\hat{r}) = \mathbf{E}^t(\mathbf{r}) r e^{j2\pi r/\lambda}$ is the far-field function of the DUT when transmitting while excited with the current I_t at its terminals, and r is the far-field distance where the electric field $\mathbf{E}^t$ is defined. The phase reference point for $\mathbf{G}^t(\hat{r})$ is also $\mathbf{p}_{\text{ref}}$.

It can be shown [23, Eq.(5.42)] that, for the receiving DUT, the relation between b_r^{dut} and V_{oc} is

$$b_r^{\text{dut}} = \frac{V_{\text{oc}} \sqrt{Z_0}}{Z_{\text{in}}^{\text{dut}} + Z_0} = -\frac{2j\lambda}{\eta I_t} \frac{\sqrt{Z_0}}{Z_{\text{in}}^{\text{dut}} + Z_0} G_p^t E_p^{\text{pw}}. \quad (10)$$

Substituting (1) and (10) into (8), we get

$$b^{\text{meas}} = -\frac{1}{a} \frac{j\lambda}{\eta} G_p^t E_p^{\text{pw}}. \quad (11)$$

The p -polarized component of the EIRP pattern in direction $\hat{r}$ is obtained by substituting (4) into (11):

$$G_p(\hat{r}) = \frac{2j\sqrt{\eta\pi}}{\lambda E_p^{\text{pw}}(-\hat{r}, \mathbf{p}_{\text{ref}})} a b^{\text{meas}}. \quad (12)$$

Eq. (12) states that, to measure $G_p(\hat{r})$ of the DUT via b^{meas} , the chamber array needs to employ weights $\mathbf{w}$ that would create a p -polarized PW E_p^{pw} in the transmitting situation when excited by a , which is incident on the DUT from the $\hat{r}$ direction. In the calibration procedure that follows, the weights are normalized such that $E_p^{\text{pw}} = 1$ and $a = 1$. The point $\mathbf{p}_{\text{ref}}$ is also the phase reference point for the reconstructed $\mathbf{G}(\hat{r})$. To reconstruct $\mathbf{G}(\hat{r})$ for multiple directions $\hat{r}$, multiple corresponding PWs must be synthesized. This is done during the chamber calibration procedure described below.

Thus, once the chamber has been calibrated, a single coherent snapshot of the CAA port signals is sufficient to reconstruct the DUT's complex EIRP pattern over all synthesized directions.

D. Chamber calibration procedure

The synthesis of the PW E_p^{pw} (i.e. computing the excitation vector $\mathbf{w}$) in Eq. (12) can be performed by utilizing e.g. the LCMV beamformer as explained in [17], which is recapped below. The beamforming is repeated for each direction $\hat{\mathbf{r}}$ and polarization p . The beamforming procedure requires that the so-called *embedded element fields* (EEFs) in the TZ are simulated or measured³ for each element of the CAA. The EEF of n -th CAA element is a vector E-field distribution in the TZ volume when the excitation coefficient for this element is unity (i.e. $w_n = 1$) and all others are zero.

If in (12) we assume $a = 1$, then from Fig. 3 it follows that during EEF simulation/measurement the corresponding CAA element should be excited with the incident wave $a_n^{\text{ca}} = 1/\sqrt{N_{\text{el}}}$, to satisfy the power conservation in the ideal power splitter. After all EEFs are known, the vector of excitation coefficients $\mathbf{w}_p(\hat{\mathbf{r}})$ is calculated for each direction $\hat{\mathbf{r}}$ and polarization p of the synthesized PW.

For an arbitrary $\hat{\mathbf{r}}$ and p , the LCMV beamformer coefficients $\mathbf{w}$ can be found by solving [17]

$$\begin{cases} \min_{\mathbf{w}} & \mathbf{w}^H \mathbf{w} \\ \text{s.t.} & \mathbf{w}^H \mathbf{H} = \mathbf{g}^H \end{cases}, \quad (13)$$

where we minimize the total array excitation power, subject to linear equality constraints on the E-field values inside the TZ. The desired E-field at the M_{tz} constraint points is specified in the vector $\mathbf{g} = [\mathbf{g}_1, \dots, \mathbf{g}_m, \dots, \mathbf{g}_{M_{\text{tz}}}]^T \in \mathbb{C}^{2M_{\text{tz}} \times 1}$, where $\mathbf{g}_m = [g_m^{\text{co}}, g_m^{\text{xp}}]$ is the co- and cross-polar components of the desired E-field at m -th point of the TZ arranged as a row vector, and T denotes the transposition operator. The desired field $\mathbf{g}$ corresponds to the E-field values that a p -polarized plane wave, incident from direction $\hat{\mathbf{r}}$ and phase-referenced to the point $\mathbf{p}_{\text{ref}}$, would produce at the constraint points. The transfer matrix $\mathbf{H} = [\mathbf{H}_1, \dots, \mathbf{H}_n, \dots, \mathbf{H}_{N_{\text{el}}}]^T \in \mathbb{C}^{N_{\text{el}} \times 2M_{\text{tz}}}$ is a set of N_{el} EEFs defined at the same M_{tz} points inside the TZ, where the field constraints $\mathbf{g}$ are defined. The n -th EEF $\mathbf{H}_n$ is arranged as a column vector $\mathbf{H}_n = [E_{n,1}^{\text{co}}, E_{n,1}^{\text{xp}}, \dots, E_{n,m}^{\text{co}}, E_{n,m}^{\text{xp}}, \dots, E_{n,M_{\text{tz}}}^{\text{co}}, E_{n,M_{\text{tz}}}^{\text{xp}}]^T$, where $E_{n,m}^{\text{co}}$ and $E_{n,m}^{\text{xp}}$ are the co- and cross-polar components of the E-field at m -th point of the TZ when only n -th array element is excited with $a_n^{\text{ca}} = 1/\sqrt{N_{\text{el}}}$ and the excitation of all other elements is set to zero (terminated in invariant loads of the backend circuitry, not necessary ideally matched).

The solution to (13) is [24, p. 587]

$$\mathbf{w}^H = \mathbf{g}^H \mathbf{Q}_{\text{reg}}^{-1} \mathbf{H}^H, \quad (14)$$

where $\mathbf{Q}_{\text{reg}} = \mathbf{Q} + (\kappa - 1) \text{diag}(\mathbf{Q})$ is the regularized version of the matrix $\mathbf{Q} = \mathbf{H}^H \mathbf{H}$ with regularization factor $\kappa > 1$.

³In practice, EEFs can be measured by means of e.g. a planar scanner with a dual-polarized probe antenna. The probe response will be a sum of responses to multiple PWs corresponding to the WG modal distribution weighted by the probe radiation pattern. Therefore, to obtain the vector E-field of EEFs, a dedicated probe-correction technique must be applied as presented in Sec. IV.

Regularization is required because $\mathbf{Q}$ becomes ill-conditioned when the number of field equality constraints in the TZ exceeds the number of degrees of freedom of the beamformer. Moreover, inside a WG, it is extremely difficult to realize a pure plane wave in the TZ, as required by the linear constraint in (13). This typically leads to significant spillover and large values of $\mathbf{w}$, implying that CAA elements would require high excitation power. The factor κ is used to relax the constraints on the TZ field and to allow for a trade-off between the TZ quality and the power intercepted by it. This relaxation technique is also known as diagonal loading and is often used to improve the robustness of beamformers [24]. In the results presented in Sec. III, $\kappa = 1.1$ was used.

The above procedure of finding $\mathbf{w}_p(\hat{\mathbf{r}})$ for a range of $\hat{\mathbf{r}}$ and all p , including the choices $E_p^{\text{pw}}(\mathbf{p}_{\text{ref}}) = 1$ and $a = 1$, is called the *chamber calibration*.

Although $p \in \{\text{co}, \text{xp}\}$ is used throughout this paper, the formulation is not restricted to linearly polarized DUTs. Since the reconstructed quantity is the complex vector EIRP pattern, both magnitude and phase of the Ludwig-3 co- and cross-polarized components are recovered. Hence, RHCP and LHCP components can be obtained by a subsequent polarization-basis transformation. Alternatively, the calibration can be formulated directly using RHCP and LHCP reference plane waves, so that the reconstructed quantities are the CP components of the DUT pattern. The latter approach was used in the numerical validation of CP antennas in Sec. III and measurements in Sec. V-B.

It should be noted that under ideal conditions the chamber calibration procedure is performed only once. However, in practice, the transfer function of the RF chain between DAC/ADC and CAA elements may drift over time, and therefore the calibration may need to be repeated periodically. Propagation of this drift to the EIRP pattern measurement uncertainty is out of scope of this paper. Nevertheless, in [17] we show how the error in beamformer coefficients $\mathbf{w}$ (which may result from uncertainties in EEF measurements or the above-mentioned drift) affects the synthesized plane wave quality⁴, which in turn propagates to the pattern measurement uncertainty [8], [25].

E. DUT pattern measurement

Measurement of the EIRP pattern of the DUT $\mathbf{G}(\hat{\mathbf{r}})$ is performed in the following steps; see Fig. 2.

First, the transmitting DUT is placed in the TZ volume in which the unit PWs were synthesized during the calibration procedure. Next, the radiation field is sampled coherently by the CAA elements, the outgoing waves of which are denoted as $\mathbf{b}^{\text{ca}} = [b_1^{\text{ca}}, \dots, b_{N_{\text{el}}}^{\text{ca}}]^T$. The b-wave at the CAA beamformer output is calculated as

$$b_p^{\text{meas}}(\hat{\mathbf{r}}) = \frac{[\mathbf{w}_p(\hat{\mathbf{r}})]^T \mathbf{b}^{\text{ca}}}{\sqrt{N_{\text{el}}}}. \quad (15)$$

⁴The plane wave quality is quantified by the E-field peak-to-peak amplitude variation and peak-to-peak phase deviation from the ideal-plane-wave phase within the TZ volume, see [17] for details.

Substituting (15) into (12) and accounting for the choices $E_p^{\text{pw}}(\mathbf{p}_{\text{ref}}) = 1$ and $a = 1$ during the chamber calibration, the EIRP pattern is calculated as

$$G_p(\hat{\mathbf{r}}) = \frac{2j}{\lambda} \sqrt{\frac{\eta\pi}{N_{\text{el}}}} [\mathbf{w}_p(\hat{\mathbf{r}})]^T \mathbf{b}^{\text{ca}}. \quad (16)$$

It is seen from Eq. (16) that after a one-time chamber calibration and performing a single snapshot measurement of the b-waves from the CAA (no DUT movement), the DUT's EIRP pattern can be reconstructed in the directions in which the PWs were synthesized during the calibration.

III. NUMERICAL VALIDATION

In this section we analyze the reconstruction accuracy of the EIRP pattern of a DUT antenna by comparing $\mathbf{G}(\hat{\mathbf{r}}) = G_{\text{co}}(\hat{\mathbf{r}})\hat{\mathbf{c}}\hat{\mathbf{o}} + G_{\text{xp}}(\hat{\mathbf{r}})\hat{\mathbf{x}}\hat{\mathbf{p}}$ from Eq. (16) in the hybrid chamber, with $\mathbf{G}^{\text{ref}}(\hat{\mathbf{r}})$ obtained for the same DUT in free space.

To validate the EIRP pattern reconstruction, the hybrid chamber calibration and measurement setup were simulated in CST Studio Suite [26] using the time-domain solver. With reference to Fig. 4(a), the WG size is $(W, H, L) = (1, 1.25, 1.74) \text{ m} = (11.67, 14.6, 20.3) \lambda @ 3.5 \text{ GHz}$, which supports 1070 propagating modes at this frequency. The CAA consists of $16 \times 20 = 320$ dual-polarized bowtie antennas with inter-element spacing of 0.7λ . The motivation for such a CAA layout can be found in [17]. Furthermore, a 9×9 array of single-polarized bowtie antennas with inter-element spacing of 0.5λ located 0.25λ above a square ground plane has been chosen as the DUT.

Such an array can represent a DUT with low-gain, high-gain, as well as a DUT realizing multiply scanned beams with different side-lobe levels, depending on the element excitation vector $\mathbf{a}^{\text{dut}} = [a_1^{\text{dut}}, \dots, a_L^{\text{dut}}]^T$, where $L = 81$ is the total number of elements in the DUT array. For the results presented below, we use a simple beamformer in which the excitation coefficient of the l -th element is computed as $a_l^{\text{dut}} = A_l e^{-j\mathbf{k} \cdot \mathbf{r}_l \cdot \hat{\mathbf{r}}^{\text{beam}}}$, where $\mathbf{r}_l$ is the position of the l -th element in the array; $\hat{\mathbf{r}}^{\text{beam}} = [\sin \theta \cos \phi, \sin \theta \sin \phi, \cos \theta]$ is the selected beam scan direction; and A_l is a magnitude coefficient which realizes the magnitude taper function in the array aperture. The rotationally-symmetric Gaussian taper function is used and defined by the taper value T at the edge of the array. In this way, a low-gain DUT is emulated by setting T to a very low level (e.g. -1000 dB), effectively exciting only the central element, and a high-gain DUT is emulated by setting $T = 0 \text{ dB}$, i.e. uniformly exciting the whole array.

The b-waves used in (16) are calculated as $\mathbf{b}^{\text{ca}} = \mathbf{S}^{\text{ca,dut}} \mathbf{a}^{\text{dut}}$, where $\mathbf{S}^{\text{ca,dut}}$ is obtained from the CST model in Fig. 4(a).

To obtain the CAA weight coefficients $\mathbf{w}_p(\hat{\mathbf{r}})$, the calibration procedure is simulated by successively exciting the CAA elements with $a_n^{\text{ca}} = 1/\sqrt{N_{\text{el}}}$, computing the EEFs at TZ points in the absence of the DUT, and applying the LCMV beamformer as explained in Sec. II-D. Changing the direction $\hat{\mathbf{r}}$ and polarization p , the EIRP pattern of the DUT is reconstructed using Eq. (16). The chosen set of $\hat{\mathbf{r}}$ values defines the angular grid for the reconstructed pattern. In the results presented below, we swept θ from 0° to 80° in steps of 1° , and ϕ from 0° to 360° in steps of 2.5° (the angles

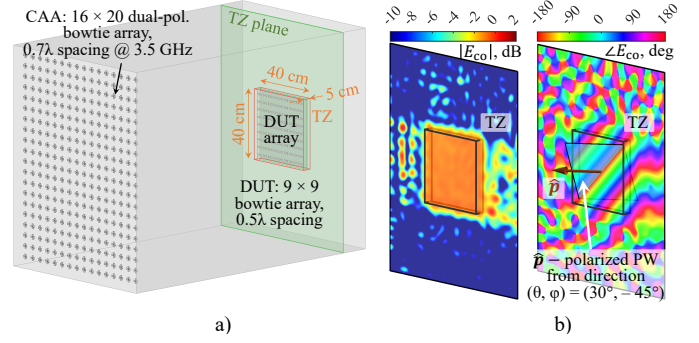


Fig. 4. (a) CST model emulating the EIRP pattern measurement. The TZ of size $40 \times 40 \times 5 \text{ cm}^3$ is denoted by a red box, inside which the DUT array is placed. (b) Magnitude and phase of the E-field co-polar component in the TZ plane when a plane wave from direction $(\theta, \phi) = (30^\circ, -45^\circ)$ is synthesized in the TZ. The simulation was performed at 3.5 GHz .

are defined in the TZ coordinate system, see Fig. 1). The maximum θ is limited to the maximum angle of incidence of the modal plane waves on the TZ, and for the considered WG dimensions and frequency is about $60^\circ \dots 80^\circ$ (depending on ϕ), see directions of the dot markers in Fig. 11(b).

To evaluate the accuracy of the EIRP pattern reconstruction, the same DUT array has been simulated in free space. The reference EIRP pattern is then calculated as $\mathbf{G}^{\text{ref}}(\hat{\mathbf{r}}) = \sum_{l=1}^L \mathbf{G}_{r,l}^{\text{dut}}(\hat{\mathbf{r}}) a_l^{\text{dut}}$, where $\mathbf{G}_{r,l}^{\text{dut}}(\hat{\mathbf{r}})$ is the realized gain embedded element pattern of the l -th DUT element. Then, we can define the relative error between the reconstructed and the reference patterns in two ways:

$$\Delta_{\text{cmplx}}(\hat{\mathbf{r}}) = |\mathbf{G}(\hat{\mathbf{r}}) - \mathbf{G}^{\text{ref}}(\hat{\mathbf{r}})| / |\mathbf{G}^{\text{ref}}(\hat{\mathbf{r}})|_{\text{max}}, \quad (17)$$

$$\Delta_{\text{mag}}(\hat{\mathbf{r}}) = \left| |\mathbf{G}(\hat{\mathbf{r}})| - |\mathbf{G}^{\text{ref}}(\hat{\mathbf{r}})| \right| / |\mathbf{G}^{\text{ref}}(\hat{\mathbf{r}})|_{\text{max}}, \quad (18)$$

where Δ_{cmplx} defines the error in the complex pattern, and Δ_{mag} shows the error in EIRP values.

The results for a low-gain DUT ($T = -1000 \text{ dB}$) and for a high-gain DUT ($T = 0 \text{ dB}$) are shown in Fig. 5 and Fig. 6, respectively. The following observations can be made. First, the error between the reconstructed and reference patterns is below -20 dB for most angles and all DUTs. Second, the accuracy is slightly worse for $\theta > 60^\circ$ which is expected because there are fewer modal PWs with such angle of arrival (see black dots in Fig. 11(b)). Third, the reconstruction accuracy is better for high-gain antennas. A possible explanation for this is that high-gain antennas have larger aperture and are less affected by inaccuracies of the synthesized plane waves [8]. This is also seen in Fig. 7, which shows the relative error, maximum and mean over directions (up to $\theta = 60^\circ$), as a function of the DUT gain realizing on-axis beam [Fig. 7(a)] or scanning in direction $(\theta, \phi) = (30^\circ, 45^\circ)$. Different gains are realized by varying the excitation taper T in the array aperture, which is also shown on the top axis. One can also note that the error level is very similar for the broadside and scanning beams, proving a high reconstruction accuracy for wide range of DUTs.

To give a more global view on the accuracy of the EIRP reconstruction, Fig. 8 shows the reconstructed and reference

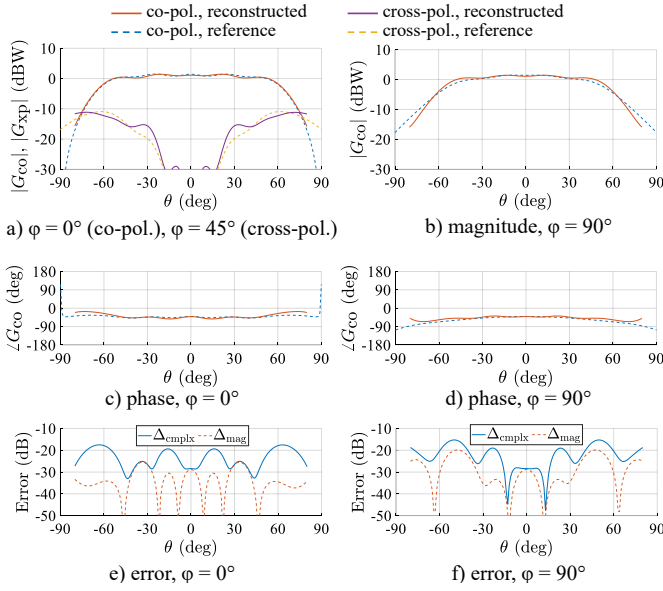


Fig. 5. Comparison of the reconstructed using Eq. (16) EIRP far-field pattern with the reference one simulated when DUT is in free space. Only the central element of the DUT array is excited with available power 0.5 W, emulating a low-gain antenna. The realized gain of the DUT is 4.5 dBi in this case. Sub-figures (a,b) and (c,d) show the pattern magnitude and phase, respectively; (e,f) the error between the reconstructed and the reference patterns, taken as complex values [Eq. (17)] and magnitude only [Eq. (18)].

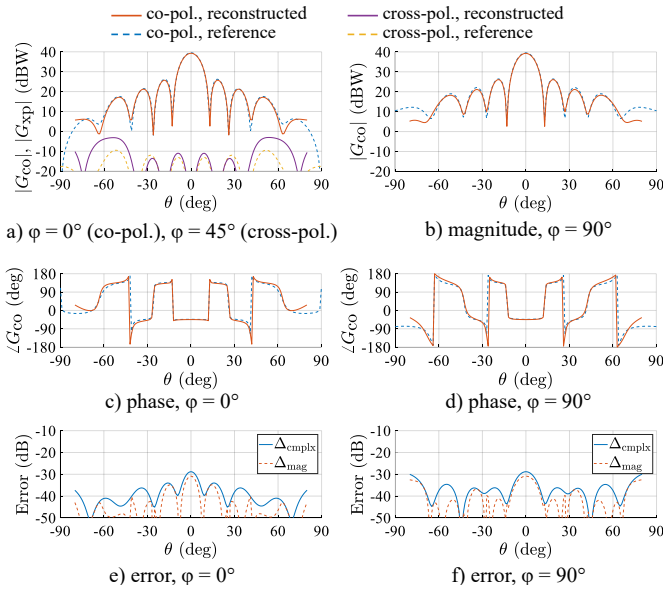


Fig. 6. Similar set of figures as in Fig. 5 but for uniformly excited DUT array emulating a high-gain antenna. The realized gain of the DUT is 23.4 dBi in this case, and each port is excited with 0.5 W.

2D EIRP patterns and the relative error between them for a low-gain and high-gain DUT scanning at $(\theta, \phi) = (0^\circ, 0^\circ)$ and $(50^\circ, 80^\circ)$. Again, the accuracy for a high-gain DUT is better despite it scans in the direction with the worst accuracy for the low-gain DUT. When the high-gain DUT scans in other directions, the relative EIRP error Δ_{mag} is below -30 dB for all (θ, ϕ) .

It is worth mentioning that CST time-domain solver was used for the simulations with a stopping criterion requiring

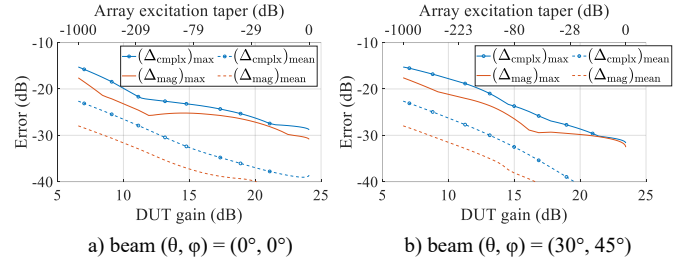


Fig. 7. Relative error between the reconstructed and reference EIRP patterns: (a) for the broadside beam of DUT, and (b) for the beam scanning in direction $(\theta, \phi) = (30^\circ, 45^\circ)$.

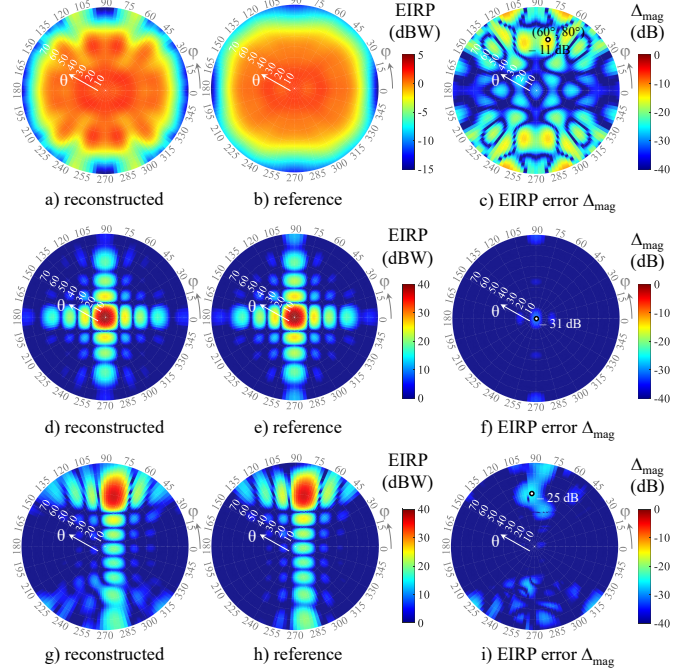


Fig. 8. The reconstructed and reference EIRP patterns and the relative error between them: (a-c) low-gain DUT; (d-f) high-gain DUT scanning at $(\theta, \phi) = (0^\circ, 0^\circ)$; (g-i) high-gain DUT scanning at $(\theta, \phi) = (50^\circ, 80^\circ)$.

the residual energy in the system below -30 dB with respect to its maximum over time. Therefore, the errors shown above below that level may be attributed to the simulation method, not to the operation of the chamber itself.

Another important note is that the above results are presented at a single frequency 3.5 GHz. However, our previous study in [17] shows that a rich spectrum of PWs can be synthesized by a similar chamber (and therefore enables EIRP pattern reconstruction) up to ≈ 4.5 GHz. This frequency upper bound is limited by the CAA inter-element spacing that should not exceed $\approx 0.8\lambda \dots 0.9\lambda$. The lower frequency bound is limited by the minimum number of propagating WG modes. As will be shown in Sec. V, the DUT pattern is reconstructed accurately at $f = 0.9$ GHz, which corresponds to only 70 propagating WG modes.

In the results presented above, linearly polarized plane waves were synthesized according to the Ludwig-3 definition, and the corresponding Ludwig-3 co- and cross-polarized components were reconstructed. However, the proposed methodol-

ogy is not limited to a linear polarization basis. Fig. 10 shows the reconstructed right-hand circularly polarized (RHC) and left-hand circularly polarized (LHC) components of an RHC-polarized 5×5 uniformly excited DUT array. In this case, to directly reconstruct the RHC/LHC components of the DUT pattern, the reference plane waves $\mathbf{g}$ in (13) are defined with RHC/LHC polarization. The figure demonstrates error levels similar to those obtained for the linearly polarized DUTs. The low values of Δ_{cmplx} further indicate that the phase of the RHC component is also accurately reconstructed.

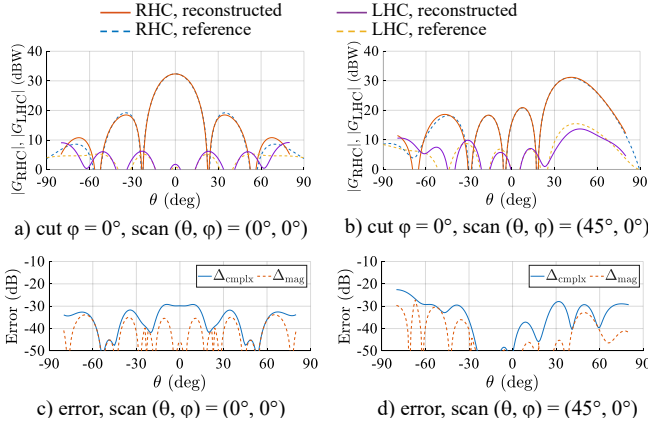


Fig. 9. Similar set of figures as in Fig. 6 but for circular-polarized 5×5 array, scanning in (a,c) broadside, and (b,d) $(\theta, \phi) = (45^\circ, 0^\circ)$. The DUT is a uniformly excited 5×5 RHCP array antenna with the phase gradient corresponding to the scan direction. The realized gain of the DUT is 18.4 dBi and 17.1 dBi for the broadside and scanned cases, respectively, and each port is excited with 1 W.

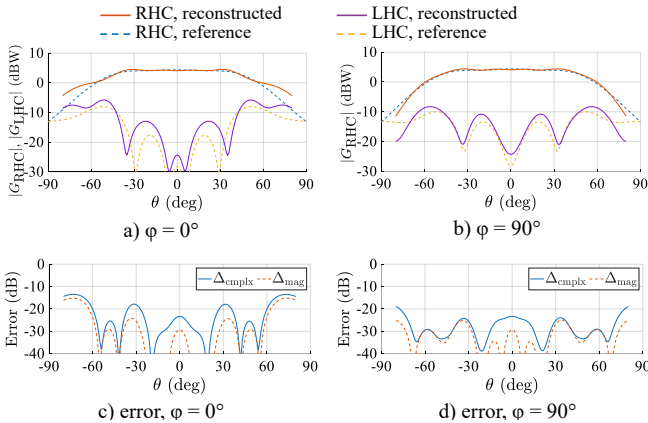


Fig. 10. Relative error between the reconstructed and reference EIRP patterns for the circular-polarized 5×5 array when only central element is excited with 1 W emulating a low-gain DUT. The realized gain of the DUT is 4.4 dBi.

IV. PROBE CORRECTION TECHNIQUE

Unlike standard FFT-based probe correction for planar near-field scans in free space, the proposed method uses the set of propagating WG modes – each represented as four coupled plane waves – as the decomposition basis, which is essential to resolve the denser modal spectrum in the overmoded WG.

As explained in Sec. II-D, to enable directional measurements in the hybrid chamber, knowledge of the vector E-field distributions in the TZ created by each individual CAA

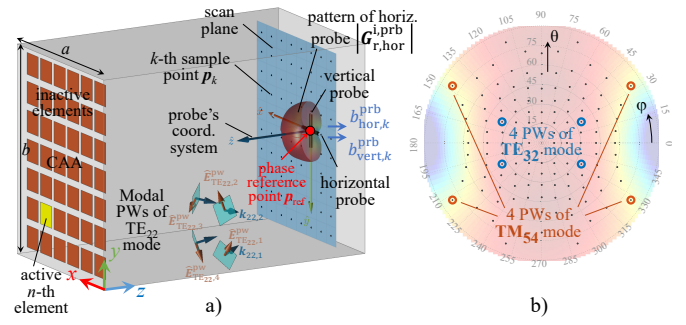


Fig. 11. On the calibration measurement and the probe correction: (a) visualization of the calibration measurement of a single EEF; in addition, four modal PWs for TE₂₂ mode are shown; (b) the modal PW directions mapped on top of the radiation pattern of the horizontal probe antenna.

element (i.e. EEFs) is required. In practice, EEFs can be measured using, e.g. a planar scanner with a dual-polarized probe, as shown in Fig. 11. Since multiple modes are typically excited in such an overmoded WG, many plane waves (PW) simultaneously impinge on the probe, which are weighted by the probe's radiation patterns and summed to obtain the measured probe's response. Moreover, the four plane waves corresponding to each WG mode have the same amplitude and phase (at their reference point). Therefore, a dedicated probe correction technique is required. In the standard probe correction technique for planar scanners in free space [3], [4], the FFT from the probe response is employed. If we use the same technique for the scan of the WG cross-section of size $W \times H$, the resolution in the spectral (plane wave) domain would be $\Delta k_x = 2\pi/W$ and $\Delta k_y = 2\pi/H$. However, the wavenumbers for the WG modes are $k_{x,m} = m\pi/W$ and $k_{y,n} = n\pi/H$, meaning that $\Delta k_x = \pi/W$ and $\Delta k_y = \pi/H$. Therefore, in the WG there are four times as many PWs than can be obtained using the FFT. To perform a probe correction in the WG, the WG modes (i.e. sets of four PWs) can be taken as the probe response decomposition basis, instead of individual PWs used in standard free-space probe correction techniques.

The core idea for reconstructing the E-field in the TZ from the measured probe responses at each sampling point of the scan plane lies in solving a system of linear equations (SLE) for the propagating WG mode coefficients. Knowing the latter, the E-field at any point close to the scan plane can be computed. Each step is detailed in the following sub-sections.

A. Probe response to a plane wave

For the EEF reconstruction from the measured probe responses, detailed in Sec. IV-C, knowledge of the probe response to a unit PW incident from an arbitrary direction is required. We define the *probe response* as 2-element b-wave vector $\mathbf{b}^{\text{prb}} = [b_{\text{hor}}^{\text{prb}}, b_{\text{vert}}^{\text{prb}}]^T$, where $b_{\text{hor}}^{\text{prb}}$ and $b_{\text{vert}}^{\text{prb}}$ are outgoing power waves from the horizontally and vertically polarized antennas of the probe, respectively. Then, when a plane wave $\mathbf{E}^{\text{pw}}$ is incident on the probe antennas from direction $\hat{\mathbf{r}}$, the probe response is (see Appendix for the derivation)

$$\mathbf{b}_{\text{pw}}^{\text{prb}}(\hat{\mathbf{r}}) = -\frac{j\lambda}{2\sqrt{\eta\pi}} \begin{bmatrix} \mathbf{G}_{\text{r,hor}}^{\text{i,prb}}(\hat{\mathbf{r}}) \cdot \mathbf{E}^{\text{pw}}(-\hat{\mathbf{r}}, \mathbf{p}_{\text{ref}}) \\ \mathbf{G}_{\text{r,vert}}^{\text{i,prb}}(\hat{\mathbf{r}}) \cdot \mathbf{E}^{\text{pw}}(-\hat{\mathbf{r}}, \mathbf{p}_{\text{ref}}) \end{bmatrix}, \quad (19)$$

where $\mathbf{G}_{r,hor}^{i,prb}(\hat{\mathbf{r}})$ and $\mathbf{G}_{r,vert}^{i,prb}(\hat{\mathbf{r}})$ are embedded realized gain patterns of the horizontal and vertical probe antennas, respectively, measured under loaded conditions in an anechoic chamber. It is noted that the probe's gain patterns, as well as the propagation direction $\hat{\mathbf{r}}$, are defined in the "probe's coordinate system" [see Fig. 11(a)], which moves during the scan. Also, the E-field vector of the PW is taken at the phase reference point of the gain patterns, which is in the origin of the probe's coordinate system.

B. Probe response to propagating WG modes

Each propagating mode in the rectangular waveguide with cross-section size $a \times b$ can be decomposed into four plane waves with the wavevectors

$$\mathbf{k}_{mn} = [\pm k_{x,m}, \pm k_{y,n}, k_{z,mn}]^T, \quad (20)$$

where $k_{x,m} = m\pi/a$; $k_{y,n} = n\pi/b$; $k_{z,mn} = \sqrt{k^2 - k_{x,m}^2 - k_{y,n}^2}$ (for forward-propagating modes); $m, n = 0, 1, 2, \dots$ (excluding $(m, n) = (0, 0)$) for TE-modes, and $m, n = 1, 2, \dots$ for TM-modes. Each combination of "+" and "-" in (20) corresponds to q -th PW the mode consists of, where $q = 1, 2, 3, 4$, and is denoted as $\mathbf{k}_{mn,q}$.

For a TE_{mn} mode, the E-field vector from q -th unit PW at an arbitrary point $\mathbf{p}$ is

$$\hat{\mathbf{E}}_{\text{TE}_{mn,q}}^{\text{pw}}(\mathbf{p}) = \frac{\mathbf{k}_{mn,q} \times \hat{\mathbf{z}}}{\|\mathbf{k}_{mn,q} \times \hat{\mathbf{z}}\|} e^{-j\mathbf{k}_{mn,q} \cdot \mathbf{p}}. \quad (21)$$

Similarly, for a TM_{mn} mode, since H-field vector is transverse ($\mathbf{H}_{mn}^{\text{pw}} \propto \mathbf{k}_{mn} \times \hat{\mathbf{z}}$), the E-field from a q -th unit PW can be expressed as

$$\hat{\mathbf{E}}_{\text{TM}_{mn,q}}^{\text{pw}}(\mathbf{p}) = \frac{(\mathbf{k}_{mn,q} \times \hat{\mathbf{z}}) \times \mathbf{k}_{mn,q}}{\|(\mathbf{k}_{mn,q} \times \hat{\mathbf{z}}) \times \mathbf{k}_{mn,q}\|} e^{-j\mathbf{k}_{mn,q} \cdot \mathbf{p}}. \quad (22)$$

The probe response to TE_{mn} mode is then calculated as sum of responses (19) to each PW obtained using (21) when the probe is located at the sampling point $\mathbf{p}$:

$$\mathbf{b}_{\text{TE}_{mn}}^{\text{prb}}(\mathbf{p}) = \frac{-j\lambda}{2\sqrt{\eta\pi}} \sum_{q=1}^4 \left[\mathbf{G}_{r,hor}^{i,prb}(-\hat{\mathbf{k}}'_{mn,q}) \cdot \hat{\mathbf{E}}_{\text{TE}_{mn,q}}^{\text{pw}}(\mathbf{p}) \right] + \left[\mathbf{G}_{r,vert}^{i,prb}(-\hat{\mathbf{k}}'_{mn,q}) \cdot \hat{\mathbf{E}}_{\text{TE}_{mn,q}}^{\text{pw}}(\mathbf{p}) \right], \quad (23)$$

where $\hat{\mathbf{k}}'_{mn,q} = \mathbf{T}\mathbf{k}_{mn,q}/\|\mathbf{k}_{mn,q}\|$, $\hat{\mathbf{E}}_{\text{TE}_{mn,q}}^{\text{pw}} = \mathbf{T}\hat{\mathbf{E}}_{\text{TE}_{mn,q}}^{\text{pw}}$, and $\mathbf{T}$ is the vector rotation matrix from the global to the probe's coordinate system, in which the probe patterns are defined. The directions of all modal PWs mapped on the magnitude pattern of the horizontal probe antenna are shown as black dots in Fig. 11(b), as well as the quadruples of PWs for modes TE_{32} and TM_{54} as an example.

The probe response $\mathbf{b}_{\text{TM}_{mn}}^{\text{prb}}(\mathbf{p})$ to the TM_{mn} mode is calculated similarly using PWs from (22).

C. EEF reconstruction

The EEF measurement procedure that also includes the probe correction is summarized in Algorithm 1.

The calibration measurement of n -th EEF (lines 2-6 in the algorithm) is visualized in Fig. 11. It starts from excitation of n -th CAA element with $a_n^{\text{ca}} = 1/\sqrt{N_{\text{el}}}$ (line 2). The

Algorithm 1: EEF measurement algorithm

```

1 for each CAA element  $n = 1 \dots N_{\text{el}}$  do
2   Excite the CAA element with  $a = 1/\sqrt{N_{\text{el}}}$ , see
   Sec. II-D;
3   Perform the scan: measure the probe response  $\mathbf{b}_{\text{meas}}^{\text{prb}}(\mathbf{p}_k)$ 
   at each point  $\mathbf{p}_k$  of the scan plane, see Eq. (24);
4   Build the probe response matrix  $\mathbf{B}$ , each element of
   which is the b-wave of one of the probe antennas to 4
   PWs of each WG mode at each point  $\mathbf{p}_k$ , see Eq. (25);
5   Solve system of equations (26) to obtain mode
   coefficients  $\mathbf{A}$ ;
6   Sum the modal PWs (21)-(22) weighted by
   corresponding  $\mathbf{A}$  at every point of the TZ  $\mathbf{p}_{\text{tz}}$  to obtain
   the EEF value there, i.e.  $\mathbf{E}_n^{\text{caa}}(\mathbf{p}_{\text{tz}})$ ;

```

Result: EEF distribution in the TZ $\mathbf{E}_n^{\text{caa}}(\mathbf{p}_{\text{tz}})$ for all CAA elements.

probe response $\mathbf{b}_{\text{meas}}^{\text{prb}}(\mathbf{p}_k)$ is measured with the aid of a planar scanner at every point $\mathbf{p}_k$ on the scan plane (line 3), where $k = 1 \dots K$, and K is the number of sample points in the scan plane. The measured probe response distribution is arranged in a vector as

$$\mathbf{b}_{\text{meas}}^{\text{prb}} = [b_{\text{hor}}^{\text{prb}}(\mathbf{p}_1), \dots, b_{\text{hor}}^{\text{prb}}(\mathbf{p}_K), b_{\text{vert}}^{\text{prb}}(\mathbf{p}_1), \dots, b_{\text{vert}}^{\text{prb}}(\mathbf{p}_K)]^T. \quad (24)$$

In line 4, the calculated probe response to all WG modes at every sampling point of the scan plane [Eq. (23)] is organized in a $2K \times N_{\text{modes}}$ matrix $\mathbf{B}$ as

$$\mathbf{B} = \begin{bmatrix} b_{\text{hor,TE}_{10}}^{\text{prb}}(\mathbf{p}_1) & b_{\text{hor,TE}_{01}}^{\text{prb}}(\mathbf{p}_1) & \dots & b_{\text{hor,TM}_{uv}}^{\text{prb}}(\mathbf{p}_1) \\ \vdots & \vdots & \ddots & \vdots \\ b_{\text{hor,TE}_{10}}^{\text{prb}}(\mathbf{p}_K) & b_{\text{hor,TE}_{01}}^{\text{prb}}(\mathbf{p}_K) & \dots & b_{\text{hor,TM}_{uv}}^{\text{prb}}(\mathbf{p}_K) \\ b_{\text{vert,TE}_{10}}^{\text{prb}}(\mathbf{p}_1) & b_{\text{vert,TE}_{01}}^{\text{prb}}(\mathbf{p}_1) & \dots & b_{\text{vert,TM}_{uv}}^{\text{prb}}(\mathbf{p}_1) \\ \vdots & \vdots & \ddots & \vdots \\ b_{\text{vert,TE}_{10}}^{\text{prb}}(\mathbf{p}_K) & b_{\text{vert,TE}_{01}}^{\text{prb}}(\mathbf{p}_K) & \dots & b_{\text{vert,TM}_{uv}}^{\text{prb}}(\mathbf{p}_K) \end{bmatrix}, \quad (25)$$

where N_{modes} is the total number of propagating modes; and uv denotes maximum indices for the propagating TM mode, for which $k_{z,mn}$ is a real number.

Since the number of measured probe responses typically exceeds the number of unknown propagating WG-mode coefficients, i.e., $2K > N_{\text{modes}}$, the probe-correction problem leads to an overdetermined linear system. The coefficients of the WG modes, each represented by four unit-amplitude plane waves, are therefore estimated in the least-squares sense as

$$\mathbf{A} = \arg \min_{\mathbf{x}} \|\mathbf{B}\mathbf{x} - \mathbf{b}_{\text{meas}}^{\text{prb}}\|_2^2. \quad (26)$$

If the matrix $\mathbf{B}$ is ill-conditioned, this least-squares problem can be regularized using, e.g., truncated singular-value decomposition or Tikhonov regularization.

Finally, the vector E-field at an arbitrary point $\mathbf{p}$ inside the WG is computed as a weighted sum of the modal PWs (21) and (22):

$$\mathbf{E}(\mathbf{p}) = \sum_{i=1}^{N_{\text{modes}}} A_i \sum_{q=1}^4 \hat{\mathbf{E}}_{S_i,q}^{\text{pw}}(\mathbf{p}), \quad (27)$$

where $S \in \{\text{TE}_{10}, \text{TE}_{01}, \dots, \text{TM}_{uv}\}$ is the list of all propagating WG modes in the order in which they are defined in a row of the matrix $\mathbf{B}$.

In line 6 of the algorithm, Eq. (27) is evaluated at every point of the TZ $\mathbf{p}_{\text{tz}}$ to obtain the EEF $\mathbf{E}_n^{\text{caa}}(\mathbf{p}_{\text{tz}})$ of the n -th element of the CAA.

V. MEASUREMENT RESULTS

The measurements presented in this section serve as an end-to-end experimental proof-of-concept of the proposed reconstruction method, including calibration, probe correction, coherent CAA sampling, and absolute EIRP normalization. The prototype operates at 900 MHz, whereas the full-wave validation in Sec. III was performed at 3.5 GHz. The lower measurement frequency was selected for practical reasons: it enabled manufacturing and integration of a CAA with a moderate number of elements using the available prototype chamber and measurement backend, while still providing enough propagating WG modes to demonstrate the reconstruction principle. The relatively low frequency and constrained size of the TZ for the given chamber geometry limit the maximum gain of DUT used for validation of the method (see Table I). The 3.5 GHz simulations, on the other hand, represent a more relevant 5G NR FR1 case with an electrically larger chamber, denser modal spectrum, and larger CAA, and are therefore used to evaluate higher-gain and beam-steered DUTs.

A. The hybrid chamber prototype and measurement setup

The hybrid chamber prototype first presented in [17] has been extended with a CAA that consists of 6×7 stacked-patch antennas resonating at 900 MHz and having inter-element spacing 0.49λ and 0.53λ in horizontal and vertical directions, respectively; see Fig. 12(a). The CAA backend is connected to the left side of the overmoded WG section [Fig. 12(b)]. The flat absorber in the planar scanner [Fig. 12(c)] has been replaced with the pyramidal one to significantly reduce back-scattering and improve measurement accuracy. Two orthogonal 0.4λ -dipoles are used as the dual-polarized probe.

During the calibration phase, the scanner backend is connected to the right side of the WG section and is used to measure CAA EEFs as described in Sec. II-D, including the dedicated probe-correction (Sec. IV). In this measurement, the probe-response matrix $\mathbf{B}$ used in the probe-correction step has size 308×70 , corresponding to two probe ports sampled at $K = 154$ scan points and 70 propagating WG modes at 900 MHz. The system is therefore overdetermined. The condition number of this matrix is $\text{cond}(\mathbf{B}) = 12.9$, and its normalized singular-value spectrum is shown in Fig. 13. This moderate condition number indicates that the modal-coefficient retrieval is not strongly ill-conditioned for the present experimental configuration. Consequently, no additional Tikhonov regularization or singular-value truncation was applied in the reported reconstruction; the modal coefficients were obtained using the least-squares solution of Eq. (26).

During the measurement phase, the scanner is replaced by a DUT backed with the absorbing wall; see Fig. 12(d).

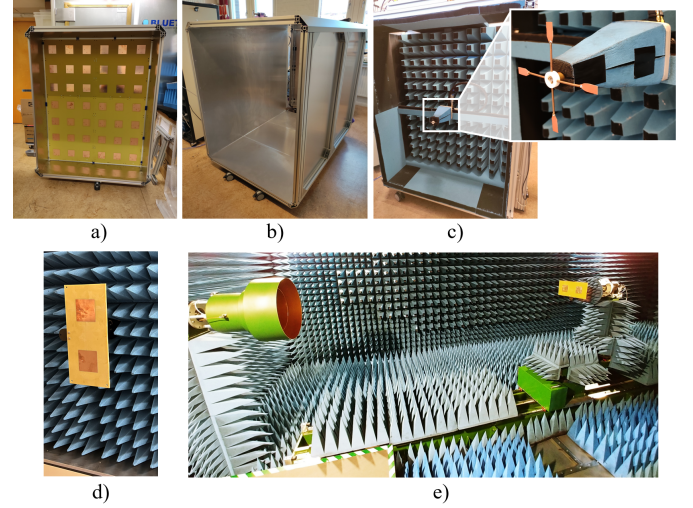


Fig. 12. The measurement setup. (a-c) Three sections of the hybrid chamber: (a) the patch CAA, (b) overmoded WG section, (c) the planar scanner. (d) Two-element patch array backed by an equal power divider used as a DUT. (e) Reference measurement setup in an AC.

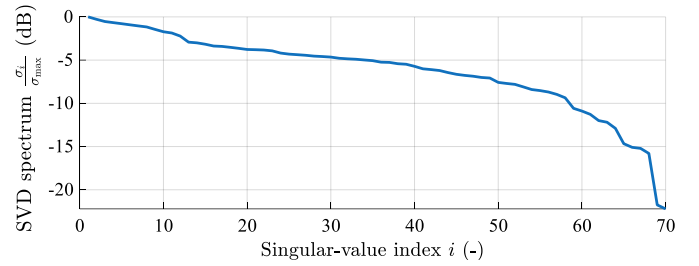


Fig. 13. Normalized singular-value spectrum of the probe-response matrix $\mathbf{B}$ used for modal coefficient retrieval in the 900 MHz measurement. The matrix size is 308×70 , corresponding to two probe ports, 154 scan points, and 70 propagating WG modes. The condition number is $\text{cond}(\mathbf{B}) = 12.9$.

Three DUTs were measured and labeled as “M1”, “M2” and “M3”: M1) a uniformly-excited linearly-polarized (LP) 2-element patch array (similar to the CAA) is used as the DUT which is mounted vertically, so that its far-field pattern is narrower in the vertical plane ($\phi = 90^\circ$); M2) a circular-polarized (CP) planar spiral antenna without ground plane; and M3) a CP planar spiral antenna with ground plane. The S-parameters between the DUT port and all CAA ports $\mathbf{S}^{\text{caa,dut}}$ were measured by a vector network analyzer (VNA)⁵, that were used to find the vector $\mathbf{b}^{\text{ca}}$. In fact, $\mathbf{b}^{\text{ca}} = \mathbf{S}^{\text{caa,dut}}$ assuming the incident power wave on the DUT is $a^{\text{dut}} = 1$, i.e. the available power at its port is $P_{\text{av}} = \frac{1}{2}|a^{\text{dut}}|^2 = 0.5$ W.

The far-field pattern of the DUT was also measured in the AC at Chalmers [27], see Fig. 12(e), and is used as the reference.

⁵The S-parameters were used to simplify the measurement setup for the proof-of-concept. However, the pattern reconstruction algorithm needs only $\mathbf{b}^{\text{ca}}$, which could be directly measured by a software-defined-radio-like backend, supporting in this way “black-box” DUTs without access to their port(s).

B. EIRP pattern measurement results

The EIRP far-field patterns of all three DUTs, directly measured in the AC and reconstructed from the hybrid chamber measurement, are shown in Figs. 14–16. To place the measured DUTs in the context of the complete validation set, Table I summarizes all DUT cases used in the paper. The simulation cases S1–S5 examine the influence of DUT aperture, gain, beam direction, and polarization basis under controlled full-wave conditions. The measurement cases M1–M3 demonstrate the complete experimental procedure in the 900 MHz prototype, including chamber calibration, probe correction, coherent CAA sampling, and EIRP normalization, using manufactured LP and CP DUTs compared with AC references.

The error metrics for representative validation cases are summarized in Table II. The table includes both simulation cases, which isolate the reconstruction behavior under controlled full-wave conditions, and measurement cases, which evaluate the complete 900 MHz prototype procedure. The table distinguishes between peak EIRP error and the maximum pointwise pattern error within the $\pm 60^\circ$ angular sector. This distinction is important because a small peak EIRP error does not necessarily imply the same accuracy over the complete pattern. It should also be noted that the reconstruction-error values in Table II should not be directly compared to the peak gain values of DUTs in Table I. The gain values are absolute realized-gain levels relative to an isotropic radiator, whereas the errors are pointwise logarithmic differences between the reconstructed and reference EIRP patterns at the same observation angle.

Comparing S1 and S2 shows the expected aperture/gain dependence: the normalized maximum error within $\pm 60^\circ$ improves from -16 dB for the low-gain central-element case to -31 dB for the uniformly excited high-gain aperture. This confirms that larger and more directive DUTs are less sensitive to local imperfections of the synthesized reciprocal plane waves. The CP cases S4 and S5 further show that the reconstruction accuracy is not limited by the use of a circular-polarization basis itself: the high-gain steered CP case gives a normalized maximum error of -32 dB, while the low-gain CP case gives -24 dB. This interpretation is further supported by the gain/beam-steering sweep S3, which indicates that the dominant trends are governed mainly by DUT aperture, directivity, and angular range.

The measured cases M1–M3 then evaluate how this reconstruction behavior carries over to the complete prototype. For M1–M3, the peak EIRP errors are 0.2, 0.4, and 0.4 dB, respectively, while the maximum pointwise EIRP-level errors within $\pm 60^\circ$ are 1, 2.4, and 1 dB. The corresponding normalized maximum errors are -22 , -17 , and -25 dB, respectively. The largest deviation of 2.4 dB occurs for M2, the spiral antenna without ground plane. This value is not negligible and is reported as the worst measured pointwise discrepancy. However, as seen in Fig. 15, this does not represent a broad degradation of the reconstructed pattern as this discrepancy is localized at large negative off-boresight angles, $-75^\circ \leq \theta \leq -55^\circ$, centered around $\theta \approx -65^\circ$. Thus, only the interval

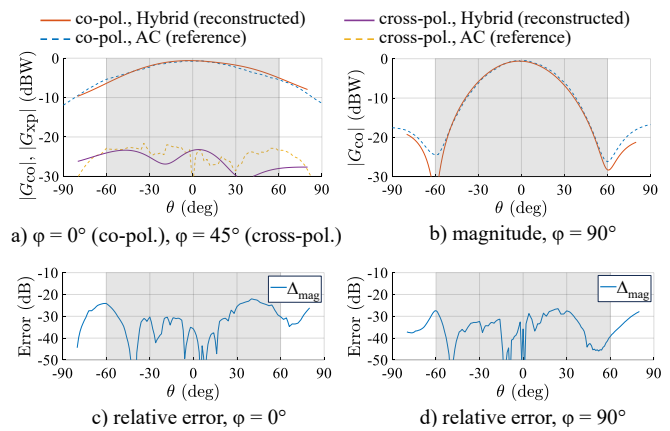


Fig. 14. The EIRP far-field pattern of the linearly-polarized 1×2 element patch DUT (case M1 in Table I).

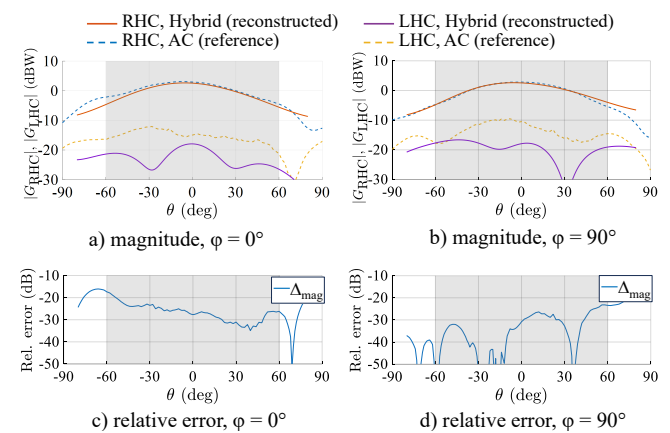


Fig. 15. The EIRP far-field pattern of the circular-polarized spiral DUT without ground plane (case M2 in Table I).

$-60^\circ \leq \theta \leq -55^\circ$ overlaps the $\pm 60^\circ$ evaluated range, i.e., the largest reported error occurs at the boundary of this range.

Furthermore, Fig. 15 shows that the AC reference pattern for M2 contains a local bump within $-75^\circ \leq \theta \leq -55^\circ$. Since the spiral antenna without a ground plane has a broad pattern with significant back radiation, it is particularly sensitive to residual AC reflections. This sensitivity is further increased at 900 MHz, close to the lower operating-frequency limit of the AC, where absorber performance and residual multipath are more critical. The localized deviation therefore likely reflects a combined effect of residual AC-reference uncertainty, the broad pattern DUT, and reconstruction uncertainty. When this localized interval is excluded, the maximum pointwise error within $\pm 60^\circ$ is reduced from 2.4 dB to 1 dB, and the corresponding normalized maximum error changes from -17 dB to -23 dB.

The measured results therefore demonstrate the complete experimental reconstruction procedure for representative low- and medium-gain LP/CP DUTs, while also showing that localized reference-chamber features, DUT back radiation, and residual scattering can dominate the worst-case pointwise pattern error.

TABLE I
SUMMARY OF DUT CASES AND TEST SCENARIOS USED TO VALIDATE THE PROPOSED PATTERN AND EIRP RECONSTRUCTION METHOD.

Case	DUT / Configuration	Validation	Freq.	Peak Gain	Pol.	Beam State	Effective TZ area	Purpose of validation
S1	9×9 bowtie element array; central element excited	Sim. vs. FS	3.5 GHz	4.5 dBi	LP	Broadside beam	Small	Low-gain
S2	9×9 bowtie element array; uniformly excited aperture	Sim. vs. FS	3.5 GHz	23.4 dBi	LP	Broadside beam	Large	High-gain
S3	9×9 bowtie element array; excitation taper varied	Sim. vs. FS	3.5 GHz	4.5–23.4 dBi	LP	Broadside and 30° steered beam	Large, medium, small	LP beam steering
S4	5×5 RHCP bowtie element array; uniformly excited	Sim. vs. FS	3.5 GHz	17–18 dBi	CP	Broadside and 45° steered beam	Large	CP high gain, beam steering
S5	5×5 RHCP bowtie element array; central element excited	Sim. vs. FS	3.5 GHz	4.4 dBi	CP	Broadside beam	Small	CP low gain
M1	1 × 2 patch element array; uniformly excited	Meas. vs. AC	900 MHz	2.3 dBi	LP	Broadside beam	Small	Experimental LP low-gain
M2	Spiral antenna without ground plane	Meas. vs. AC	900 MHz	2.9 dBi	CP	Broadside beam	Small	Experimental CP low-gain
M3	Spiral antenna with ground plane	Meas. vs. AC	900 MHz	5.0 dBi	CP	Broadside beam	Medium	Experimental CP medium gain

Notes: TZ = test zone; FS = free space (simulation); AC = anechoic chamber; LP = linear polarization; CP = circular polarization; the peak realized gain values are absolute levels in dBi relative to an isotropic radiator.

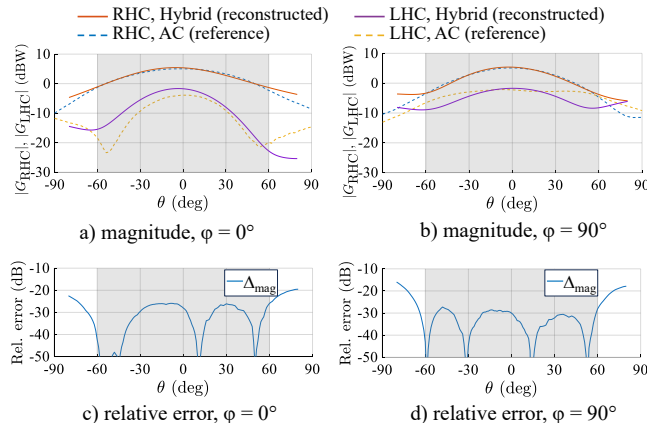


Fig. 16. The EIRP far-field pattern of the circular-polarized spiral DUT with ground plane (case M3 in Table I).

TABLE II
SUMMARY OF NORMALIZED AND ABSOLUTE EIRP PATTERN RECONSTRUCTION ERRORS FOR REPRESENTATIVE VALIDATION CASES IN TABLE I.

Case	Abs. max. error at peak	Norm. max. error at peak	Abs. max. error within ±60°	Norm. max. error within ±60°
S1	0.3 dB	−28 dB	1.1 dB	−16 dB
S2	0.2 dB	−31 dB	0.9 dB	−31 dB
S4	0.05 dB	< −30 dB	1.1 dB	−32 dB
S5	0.3 dB	−29 dB	0.6 dB	−24 dB
M1	0.2 dB	−35 dB	1 dB	−22 dB
M2	0.4 dB	−27 dB	2.4*, 1** dB	−17*, −23** dB
M3	0.4 dB	−28 dB	1 dB	−25 dB

The absolute errors in this table are pointwise EIRP-level differences in dB between reconstructed and reference patterns. For the simulated cases shown in the table, errors are relative to the corresponding free-space reference using Δ_{mag} from (18). Case S3 is a parametric gain/beam-steering sweep and is summarized in Fig. 7. For M1–M3, errors are relative to the AC measurement.

* Including the localized feature around $-75^\circ \leq \theta \leq -55^\circ$ in Fig. 15, which is discussed in Sec. V-B and is likely influenced by the AC reference measurement.

** Excluding the angular interval $-75^\circ \leq \theta \leq -55^\circ$.

VI. PRACTICAL APPLICABILITY AND USE CASES

A. DUT Generality and Polarization Basis

The validation cases in this work were selected to examine complementary aspects of the reconstruction method, includ-

ing DUT electrical aperture, gain, polarization basis, beam direction, and experimental implementation constraints. The full-wave simulations at 3.5 GHz cover low- and high-gain apertures, broadside and tilted beams, and LP/CP polarization bases. The 900 MHz prototype measurements demonstrate the complete experimental procedure, including chamber calibration, probe correction, coherent CAA sampling, and absolute EIRP normalization, using measured LP and CP DUTs compared with AC measurements as references.

The selection of measured DUTs is constrained by the electrically small TZ of the present 900 MHz prototype. This constraint implies that a substantially higher-gain LP or CP DUT would approach or exceed the validated TZ dimensions. Therefore, high-gain, narrow-beam, and electronically tilted-beam cases are evaluated in the 3.5 GHz full-wave model.

The formulation is not restricted to linearly polarized DUTs. Although Ludwig-3 co- and cross-polarized components are used in most equations, the reconstructed quantity is the complex vector EIRP pattern. Therefore, both magnitude and phase of two orthogonal polarization components are recovered. Circular-polarization components can be obtained by a subsequent polarization-basis transformation, e.g.,

$$\begin{bmatrix} G_{\text{RHCP}}(\hat{\mathbf{r}}) \\ G_{\text{LHCP}}(\hat{\mathbf{r}}) \end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix} 1 & j \\ 1 & -j \end{bmatrix} \begin{bmatrix} G_{\text{co}}(\hat{\mathbf{r}}) \\ G_{\text{xp}}(\hat{\mathbf{r}}) \end{bmatrix}, \quad (28)$$

where the exact sign convention depends on the adopted definition of RHCP and LHCP. Alternatively, the chamber calibration can be formulated directly in terms of RHCP and LHCP reference plane waves, so that the reconstructed quantities are the CP components of the DUT pattern.

B. Single-Snapshot Operation and Waveform Measurements

The term “single snapshot” in this work denotes a simultaneous coherent acquisition of all CAA receiver channels for one stationary DUT transmit state. It should not be interpreted as an infinitesimal time sample. For a narrowband stationary signal, the CAA output can be represented as $\mathbf{b}^{\text{ca}}(t) = s(t)\mathbf{b}_0^{\text{ca}} + \mathbf{n}(t)$, where $s(t)$ is a common complex waveform factor, $\mathbf{b}_0^{\text{ca}}$ is the CAA response corresponding to

the DUT radiation state (e.g. active antennas, beam direction, positioning in the TZ), and $\mathbf{n}(t)$ denotes receiver noise. A common phase variation in $s(t)$ changes only the absolute phase reference of the reconstructed complex pattern and does not distort the EIRP magnitude pattern. A common amplitude variation scales the reconstructed EIRP and must be accounted for when absolute power is required. In contrast, non-common changes of beam state, power-control state, or radiation mechanism during the coherent acquisition interval contribute to uncertainty or lead to an average over those states.

For coherently sampled waveforms, the recorded time-domain signals from all CAA channels can be transformed to the frequency domain, and the reconstruction can be applied per frequency or sufficiently narrow sub-band using the corresponding calibration weights [17]. A full wide-band implementation requires waveform synchronization and sub-band-dependent calibration and is identified as future work.

C. Calibration Effort and System Stability

The method separates chamber calibration from per-DUT acquisition. In the present prototype, the EEFs were measured sequentially using a VNA and the complete calibration required approximately one hour. This time is implementation-specific. Once the EEFs and beamforming weights are known, the per-DUT or per-DUT-state measurement consists of one coherent CAA acquisition followed by digital post-processing.

Amplitude and phase drift of the RF chains changes the synthesized reciprocal plane waves and therefore contributes to reconstruction uncertainty. In practical implementations, such drift can be monitored by periodic measurements of a reference radiator, internal calibration paths, or CAA-channel transfer-function checks. If the measured drift exceeds a specified tolerance, the affected RF-chain calibration should be updated. A standardized long-term uncertainty budget, including drift and recalibration intervals, is identified as future work.

D. Operational Frequency Range and Modal Scaling

The frequency range of operation is governed by the modal content of the overmoded WG and by the spatial sampling of the CAA. At low frequencies, fewer propagating WG modes are available, which reduces the angular diversity of the reciprocal plane-wave basis. At higher frequencies, the number of propagating modes increases and can improve angular coverage and angular resolution, but only if the CAA element spacing, number of RF channels, calibration accuracy, and channel stability scale accordingly. In the present prototype, 70 propagating WG modes are available at 900 MHz. The frequency-dependent plane-wave synthesis coverage of a similar chamber was analyzed in [17] using the concept of operational bandwidth.

Fig. 17 gives additional insight into the frequency behavior by plotting the EIRP at $(\theta, \phi) = (0^\circ, 0^\circ)$ for two measured DUTs. A quasi-periodic ripple is observed in both the reconstructed result and the AC reference, which is consistent with a standing-wave effect in the measurement environment. Since

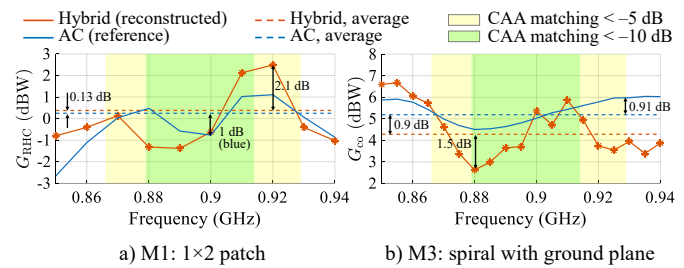


Fig. 17. The EIRP in $(\theta, \phi) = (0^\circ, 0^\circ)$ of (a) the linearly-polarized 1×2 element patch DUT (case M1 in Table I) and (b) the spiral antenna with ground plane (case M3 in Table I). The EIRP average is taken within 60 MHz bandwidth (0.87–0.93 GHz) which approximately corresponds to one period of the standing wave.

this ripple is also present in the AC reference and remains visible within the bandwidth where the chamber array is well matched, it should not be attributed to the CAA impedance mismatch effect alone. The reconstructed and reference values averaged over 0.87–0.93 GHz agree within 0.13 dB and 0.9 dB for DUT cases M1 and M3, respectively. In contrast, the maximum deviation from the corresponding average reaches 2.1 dB and 1.5 dB in the reconstructed traces, and 1.0 dB and 0.91 dB in the AC reference traces, for M1 and M3, respectively. This behavior is consistent with the stronger multi-scattering environment of the hybrid chamber and indicates that frequency-dependent ripple should be included in the practical uncertainty assessment.

From a practical perspective, this effect can be reduced by designing a CAA with low radar cross-section (RCS) and by improving termination of the WG open end behind the DUT. For applications where a sub-band-averaged EIRP metric is acceptable, a practical mitigation is to average the EIRP over approximately one ripple period, which is around 60 MHz for the measured cases.

E. Positioning Relative to Existing Antenna Range Methods

To position the proposed method relative to other antenna measurement range methods, we follow the range-selection considerations of IEEE Std 149-2021 [28]. In doing so, we distinguish between the four most relevant aspects, i.e., DUT class and electrical size, operating frequency range, accuracy-related metrics, and measurement-time scalability with the number of DUT states, because different ranges are optimized for different combinations of these requirements, and their reported accuracy metrics are not always directly equivalent.

Conventional AC and CATR methods provide well-established references for high-accuracy pattern and gain measurements. The proposed method has similar objectives, but changes how the angular samples are acquired: they are obtained digitally from one coherent CAA response vector instead of by rotating the AUT or by changing the line of sight.

IEEE Std 149-2021 (see Sec. 4.4.1, Table 1) recommends different ranges depending on frequency and AUT size. For example, CATRs are recommended for medium-to-electrically-large AUTs (with $> 5\lambda$ – 10λ aperture diameter) at frequencies above 1-2 GHz, where a conventional AC would become long.

CATRs are more common for higher-frequency and electrically larger DUTs due to physical-size constraints. The hybrid chamber targets complementary use cases, where frequencies can be below 2-GHz and DUT beamforming varies, resulting in varying effective aperture.

Another distinction concerns the performance metrics. CATRs are commonly characterized in terms of gain and sidelobe uncertainty at selected pattern levels. For example, Blalock and Fordham [29] report an expanded uncertainty of 0.64 dB for gain and 1.06 dB at a -30 dB sidelobe based on a combined uncertainty budget, whereas Gregson and Parini [30] predict approximately ± 1 dB uncertainty at a -20 dB sidelobe from computational modelling of conventional CATR quiet-zone specifications. These scalar uncertainties are not directly equivalent to the normalized complex or magnitude EIRP reconstruction errors used here to assess the complete complex-field pattern shape.

Finally, the range-selection depends on the measurement-time scalability with the number of DUT states. This aspect is central for beam-steerable arrays and active transmitters, where many excitation states, power-control states, or beam directions may need to be verified. In conventional AC or CATR ranges, the acquisition time scales with the number of angular samples, frequencies, polarizations, and DUT states, because each state generally requires a separate angular measurement sequence. In MPAC/PWG systems, part of the angular control can be performed electronically. They are in this sense closely related to the proposed method, but they typically focus on channel performance rather than patterns.

MPAC/PWG-based methods use optimally beamformed arrays and have measurement uncertainty contributions similar to compact ranges [28, Sec. 12.3.6]. The literature often targets MIMO performance metrics such as EVM and throughput, rather than deterministic pattern reconstruction. For example, Gregson and Parini report simulated main-beam gain losses ranging from 0.09 to 0.5 dB, depending on the beam-scan angle and measurement distance [31], while Tankielun et al. [32], [33] report gain deviations of 0.2 dB relative to a far-field system over 2.3–3.6 GHz.

Within this measurement landscape, the proposed hybrid chamber targets complementary applications that require compactness, motionless acquisition, and short per-DUT or per-DUT-state measurement time. The proposed method is therefore assessed using the normalized EIRP-pattern reconstruction error function with respect to free-space simulations or direct far-field AC measurements which are the most relevant references for deterministic pattern reconstruction.

VII. CONCLUSIONS

A motionless far-field radiation-pattern and EIRP measurement methodology for black-box transmitters in a compact hybrid OTA chamber has been presented and validated in simulations and measurements. By combining a power-wave formulation with reciprocity, we derived a linear relation between the coherently sampled CAA receive signals and the DUT EIRP pattern. This relation enables reconstruction of the complex vector EIRP pattern from a single coherent

snapshot of the CAA signals, without any mechanical rotation of the DUT, which is one of the main practical advantages of the proposed method compared with conventional pattern-measurement approaches.

A one-time chamber calibration procedure was introduced to define the angular sampling grid for the pattern and the absolute EIRP normalization. It involves measurement of the CAA embedded element fields (EEFs) in the test zone, which are used together with a Linearly Constrained Minimum Variance (LCMV) beamformer to synthesize unit-amplitude, polarization-controlled spectrum of plane waves. This calibration is not part of the per-DUT measurement time: once the EEFs and beamforming weights are known, the radiation pattern of each DUT or DUT beam state can be reconstructed by digital post-processing of one coherent CAA acquisition. From a resolution perspective, the angular sampling grid is entirely defined in software by the set of synthesized plane waves. For the considered $16 \times 20 \times 2$ -element CAA, hundreds of angular samples over the $\theta = \pm 70^\circ$ sector in each cut can be realized using a single coherent snapshot, which satisfies typical sampling criteria for far-field characterization of the examined DUTs. The fundamental limit on the number of directions, as well as on the maximum θ -angle, is governed by the number of CAA elements and the quality of plane-wave synthesis (and, therefore, accuracy of the pattern reconstruction).

A dedicated probe-correction technique was further developed to reconstruct vector fields in an overmoded waveguide from planar scans, accounting for all propagating modes through a modal plane-wave decomposition. This step is essential for accurate EEF estimation in the strongly multi-scattering waveguide environment. In the reported 900 MHz experiment, the probe-correction system was overdetermined and moderately conditioned, which allowed the modal coefficients to be retrieved without additional regularization.

The validation set covered different DUT gains, electrical apertures, beam directions, and polarization bases. Simulations at 3.5 GHz included LP 9×9 bowtie-array cases with gains from 4.5 to 23.4 dBi, including broadside, gain-swept, and beam-steered configurations. These results show three important trends. First, the reconstruction error is below -20 dB for most angles and DUT cases. Second, the error increases at larger observation angles, especially for $\theta > 60^\circ$, which is consistent with the reduced density of modal plane-wave directions available at the edge of the angular sector. Third, higher-gain and larger-aperture DUTs are reconstructed more accurately than low-gain DUTs because they are less sensitive to local imperfections of the synthesized plane waves. For high-gain scanned cases, the magnitude reconstruction error remains below -30 dB over the evaluated 2-D angular range. The CP simulations with 5×5 RHCP arrays further show that the same calibration and reconstruction framework can recover RHC/LHC components for both broadside and 45° steered beams, with error levels similar to the LP cases. Thus, the simulations support the method for high-gain, beam-steered, and CP scenarios within the considered full-wave chamber model.

The experimental validation in the 900 MHz prototype

chamber was performed for three manufactured DUTs: an LP 1×2 patch array with 2.3 dBi peak gain, a CP spiral antenna without ground plane with 2.9 dBi peak gain, and a CP spiral antenna with ground plane with 5.0 dBi peak gain. These cases demonstrate the complete measurement procedure for representative low- and medium-gain LP/CP DUTs in the available prototype. The measured EIRP errors at the peaks relative to the AC reference were 0.2, 0.4, and 0.4 dB for the three cases, respectively, showing that the absolute peak EIRP level can be recovered with sub-dB accuracy. The corresponding maximum pointwise EIRP errors within $\pm 60^\circ$ were 1, 2.4, and 1 dB, respectively, or -22 , -17 , and -25 dB for the normalized maximum errors. This difference between peak-level and pointwise-pattern metrics is important: the maximum pointwise error is a conservative metric and can be dominated by localized angular features. The 2.4 dB maximum error occurs for the CP spiral antenna without ground plane, which is a back-radiating low-gain DUT measured near the lower operating-frequency limit of the AC reference facility. The error is localized at large negative off-boresight angles, near the boundary of the evaluated angular sector, and is reduced to 1 dB when the affected angular interval is excluded. Thus, this case defines the most challenging measured condition in this work and helps bound the practical performance envelope of the present 900 MHz prototype.

Overall, the results show that the proposed hybrid chamber can provide compact, motionless, and fast direction-resolved EIRP-pattern measurements for a fixed DUT state after chamber calibration. The present experimental evidence supports representative low- and medium-gain LP/CP DUTs, while high-gain and beam-steered DUTs, including CP cases, are supported by the 3.5 GHz full-wave validation rather than by the present prototype measurements. The results also show that the relevant accuracy metric depends on the application: peak EIRP can be recovered with sub-dB accuracy in the measured cases, while full-pattern use cases must consider conservative pointwise errors, especially at large off-boresight angles and for back-radiating, broad-beam DUTs. The method is therefore most directly relevant to rapid OTA characterization, beam-state verification, and production-oriented comparative testing rather than reference-class metrology. Future work will address a complete uncertainty budget, long-term calibration stability, wideband waveform-based operation, and experimental validation with larger active beam-steerable DUTs.

APPENDIX

Since the scanner probe can be seen as a 2-element array (horizontally and vertically polarized antennas), let us derive the array response (outgoing power waves from each element) $\mathbf{b} = [b_1, b_2]^T$ to an incident PW $\mathbf{E}^{\text{pw}}$.

The relation between $\mathbf{b}$ and the open-circuited voltages induced at the array ports $\mathbf{V}^{\text{oc}} = [V_1^{\text{oc}}, V_2^{\text{oc}}]^T$ is [23, Eq. (5.42)]

$$\mathbf{b} = \sqrt{Z_0} (\mathbf{Z} + \mathbf{Z}_0)^{-1} \mathbf{V}^{\text{oc}}, \quad (29)$$

where Z_0 is the reference impedance; $\mathbf{Z}$ is the 2×2 input impedance matrix of the probe array; $\mathbf{Z}_0 = Z_0 \mathbb{I}$; and $\mathbb{I}$ is the identity matrix.

When Eq. (9a) is applied to an element in an array, $\mathbf{G}^t(\hat{\mathbf{r}})$ is assumed to be the *embedded* element far-field function when the element is excited with the current I_t while the port currents of the other array elements are zero (i.e. they are open-circuited), i.e.

$$\mathbf{V}^{\text{oc}} = -\frac{2j\lambda}{\eta} \begin{bmatrix} \frac{1}{I_{t,\text{hor}}} & 0 \\ 0 & \frac{1}{I_{t,\text{vert}}} \end{bmatrix} \begin{bmatrix} \mathbf{G}_1^t(\hat{\mathbf{r}}) \cdot \mathbf{E}^{\text{pw}}(-\hat{\mathbf{r}}, \mathbf{p}_{\text{ref}}) \\ \mathbf{G}_2^t(\hat{\mathbf{r}}) \cdot \mathbf{E}^{\text{pw}}(-\hat{\mathbf{r}}, \mathbf{p}_{\text{ref}}) \end{bmatrix}. \quad (30)$$

Let us introduce $\mathbf{G}_p^l(\hat{\mathbf{r}})$, $p \in \{1, 2\}$, which is p -th element embedded far-field function when this element is excited with the current 1 A, and other elements are open-circuited. Then, Eq. (30) simplifies to

$$\mathbf{V}^{\text{oc}} = -\frac{2j\lambda}{\eta} \begin{bmatrix} \mathbf{G}_1^l(\hat{\mathbf{r}}) \cdot \mathbf{E}^{\text{pw}}(-\hat{\mathbf{r}}) \\ \mathbf{G}_2^l(\hat{\mathbf{r}}) \cdot \mathbf{E}^{\text{pw}}(-\hat{\mathbf{r}}) \end{bmatrix}. \quad (31)$$

In practice, the embedded realized gain patterns $\mathbf{G}_{r,1}^i(\hat{\mathbf{r}})$ and $\mathbf{G}_{r,2}^i(\hat{\mathbf{r}})$ are typically measured when all elements are terminated with Z_0 impedance.

Similarly to $\mathbf{G}_p^l(\hat{\mathbf{r}})$, we introduce $\mathbf{G}_p^s(\hat{\mathbf{r}})$ which is p -th element embedded far-field function when this element is excited with power wave $a_p = 1$, and other elements are terminated by Z_0 . In this case, according to Eq. (4),

$$\mathbf{G}_p^s(\hat{\mathbf{r}}) = \sqrt{\frac{\eta}{4\pi}} \mathbf{G}_{r,p}^i(\hat{\mathbf{r}}). \quad (32)$$

Now, we can pose a question: “With which a -waves the terminated array elements should be excited to get i) 1 A current through port of the first element and 0 A through the second one; and ii) 0 A current through port of the first element and 1 A through the second one?”. The answer provides us with the relation between $\mathbf{G}^l = [\mathbf{G}_1^l, \mathbf{G}_2^l]^T$ and $\mathbf{G}^s = [\mathbf{G}_1^s, \mathbf{G}_2^s]^T$, i.e.

$$\mathbf{G}^l(\hat{\mathbf{r}}) = \mathbf{a} \mathbf{G}^s(\hat{\mathbf{r}}), \quad (33)$$

where

$$\mathbf{a} = \frac{1}{2\sqrt{Z_0}} (\mathbf{Z} + \mathbf{Z}_0) \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} = \frac{1}{2\sqrt{Z_0}} (\mathbf{Z} + \mathbf{Z}_0). \quad (34)$$

Substitution of (34) and (32) into (33), and then (33) into (31), yields

$$\mathbf{V}^{\text{oc}} = -\frac{j\lambda}{2\sqrt{Z_0}\sqrt{\eta\pi}} (\mathbf{Z} + \mathbf{Z}_0) \begin{bmatrix} \mathbf{G}_{r,1}^i(\hat{\mathbf{r}}) \cdot \mathbf{E}^{\text{pw}}(-\hat{\mathbf{r}}) \\ \mathbf{G}_{r,2}^i(\hat{\mathbf{r}}) \cdot \mathbf{E}^{\text{pw}}(-\hat{\mathbf{r}}) \end{bmatrix}. \quad (35)$$

Finally, substituting (35) into (29), we obtain the array response to an incident PW using the measured embedded element realized gain patterns:

$$\mathbf{b} = -\frac{j\lambda}{2\sqrt{\eta\pi}} \begin{bmatrix} \mathbf{G}_{r,1}^i(\hat{\mathbf{r}}) \cdot \mathbf{E}^{\text{pw}}(-\hat{\mathbf{r}}) \\ \mathbf{G}_{r,2}^i(\hat{\mathbf{r}}) \cdot \mathbf{E}^{\text{pw}}(-\hat{\mathbf{r}}) \end{bmatrix}. \quad (36)$$

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