

A Wideband Low-Profile Phase Correcting Metasurface to Improve Directive Radiation Characteristics of Fabry–Pérot Cavity Antennas

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Abstract—The integration of printed phase-correcting metasurfaces (PCMSs) in the near-field of Fabry–Pérot cavity antennas (FPCAs) offers a powerful means to enhance their radiation performance. This paper presents a novel PCMS design that successfully generates a uniform aperture phase for the FPCA across a wide operational bandwidth, encompassing both resonant and off-resonant regimes. We realize this by investigating that the FPCAs’ nonuniform aperture phases vary quasi-linearly with frequency, following the quasi-linear phase constants of the dominant leaky-wave mode. Thus, the unit cells with linearly varying phases (or constant time delays) are the best candidates to correct the nonuniform aperture phase of FPCA across a wide bandwidth. Consequently, we construct an ultrathin PCMS encompassing unit cells with linearly varying phases against frequencies to accurately provide the required phase shifts in a wideband. We show through simulations and measurements that the proposed FPCA with the PCMS achieves a 3 dB gain bandwidth of 35% with a peak gain of 20.1 dBi and a peak aperture efficiency of 23%. We foresee the proposed FPCA being utilized in contemporary communication systems for long-distance communication and satellite connectivity.

Index Terms—Fabry–Pérot antenna, Electromagnetic Bandgap antenna, 2D Leaky wave antenna.

I. INTRODUCTION

MODERN communication systems and imaging systems require high-gain and wideband antennas with compact sizes [1]–[6]. These antennas are essential for providing higher data rates, establishing better wireless connections, improving signal quality, or increasing the resolutions of images for imaging systems. Fabry–Pérot cavity antennas (FPCAs), a class of two-dimensional leaky-wave antennas also known as resonant cavity antennas (RCAs), are a promising solution for modern communication systems due to their ability to provide high gain with a simple planar structure, low profile, simple feeding mechanism, and low cost [7]–[10].

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Figure 1(a) shows a diagram of the FPCA, whose basic construction involves creating a cavity by loading a partially reflecting surface (PRS) on the top of a primary source backed by a perfect electric conducting (PEC) ground [11], [12] or an artificial magnetic conducting (AMC) ground [13]–[15]. The PRS could be made of single or multiple dielectric slabs [6], [16]–[19], metasurfaces [7], [10], [20], or non-transparent fully reflecting surfaces [21], [22]. The primary source can be considered as a Hertzian dipole [23], a waveguide [1], [10], or a patch antenna [18], [24]. The cavity extends in the transverse direction. The wave propagates inside the cavity with multiple reflections from the PRS and the ground plane, and simultaneously leaks in the longitudinal direction due to the partially transparent nature of the PRS. The constructive interference of the leaking waves ensures directive radiation when the cavity satisfies the Fabry–Pérot resonant condition [6], [7], [10], [11], [18], involving the reflection properties of the PRS, the ground plane, operating frequency, the cavity height, and the material inside the cavity. When the resonant condition is satisfied, the FPCA achieves a peak gain at the resonant frequency, which depreciates away from the resonant frequency [25], resulting in a narrowband operation. For example, Ref. [26] demonstrated an FPCA design achieving a peak gain of 22.6 dBi with a narrow 3 dB gain bandwidth of 1.3%. However, using bandwidth enhancement techniques such as optimized partially reflecting surfaces [27], the bandwidth of FPCA can be enhanced up to 13.5% [28] and 26% [29]. Ref. [30] investigates the FPCA with a double-layer nonuniform superstrate, which achieves peak gain and 3 dB gain bandwidth of 19.4 dBi and 8%, respectively. Furthermore, Refs. [1], [31] proposed that the gain and bandwidth of the FPCA can be compromised to have smaller footprints but with limited peak gains.

Refs. [32], [33] recently discovered that even with a satisfied *resonant condition*, an FPCA undergoes a nonuniform aperture phase or near-field phase defect, arising due to the phase delays associated with the traverse propagation of wave inside the cavity. This nonuniformity in the aperture phase reduces the effective radiation aperture and ultimately lowers the peak directivity and aperture efficiency of an FPCA. Unlike a conventional lens, the FPCA’s aperture phase cannot be derived from simple lens theory due to the phase accumulation from multiple reflections within the cavity. Thus, Ref. [32] proposes a new methodology involving phase-correcting metasurface (PCMS) in the near-field of the FPCA to correct

its nonuniform aperture phase, leading to enhanced directive radiations, with stable sidelobe levels (SLL).

Several classes of PCMSs in the form of dielectric structures [32], [34], [35], printed planar metasurfaces [33], [36]–[38], and all metallic metasurfaces [39]–[41] have been investigated in the literature. Though dielectric-based PCMSs for FPCA improve the antenna's gain, their non-planar configuration, large vertical profiles ($1.5\lambda_0 - 2.5\lambda_0$), bulky structures, and high fabrication cost hinder their implementation as a viable compact solution for modern communications systems. The class of printed planar metasurfaces resolves some of these issues by providing localized time delays through their planar apertures, but they suffer from narrow operational bandwidth (around 10%) [38]. Ref. [33] utilizes a planar PCMS composed of square patches for the FPCA and realizes a peak gain and 3dB gain bandwidth of 19.3 dBi and 6.4 %, respectively. Ref. [38] applies a particle swarm optimization algorithm to design an optimized PCMS, achieving a peak gain and 3 dB gain bandwidth of 20.8 dBi and 9.4 %, respectively. The class of PCMS based on all-metallic structures is promising for high-power applications but comes with large vertical profiles ($1.5\lambda_0 - 2\lambda_0$) and narrow 3 dB gain bandwidths (7% to 12%) [39]–[41].

A review of the literature reveals that the bandwidth of printed planar PCMS-based FPCAs is fundamentally constrained by a resonant design paradigm. In conventional approaches, unit cells are optimized to provide a uniform aperture phase at a single resonant frequency, neglecting the cavity's phase behavior elsewhere. This leads to performance degradation outside a narrow band, severely limiting data rates, even with optimization [38]. While recent studies employing non-resonant cavities have shown promise in improving gain [42], [43], their operational bandwidth remains narrow, typically around 10.6%. In this work, we address this persistent bandwidth limitation by exploiting a key characteristic: the aperture phase of an FPCA cavity exhibits a quasi-linear variation with frequency. We demonstrate that by designing a PCMS whose unit cells provide a corresponding wideband linear phase compensation, the aperture phase can be uniformly corrected across a broad spectrum. This principle enables the realization of an FPCA-PCMS system with a significantly wider relative bandwidth than previously reported, directly addressing the critical need for wider bandwidths in modern communication systems.

As an example, we design an ultrathin planar PCMS (thickness = $0.13\lambda_0$) capable of correcting the nonuniform aperture phase of the FPCA in a wide operational bandwidth from 13 GHz to 18.50 GHz. Before designing the proposed PCMS, we first investigate the dispersion behavior of the dominant leaky-wave mode inside the conventional FPCA cavity. We apply the transverse resonance method and analyze the behavior of the quasi-TEM, TE, and TM leaky-wave modes. We note that the transversal propagation constants of the dominant leaky wave inside the cavity possess quasi-linearly varying phase shifts with respect to frequency. Hence, the aperture of the conventional FPCA also exhibits non-uniform phases, which vary quasi-linearly versus frequency. We envision that these nonuniformities can be rectified in a wideband by implement-

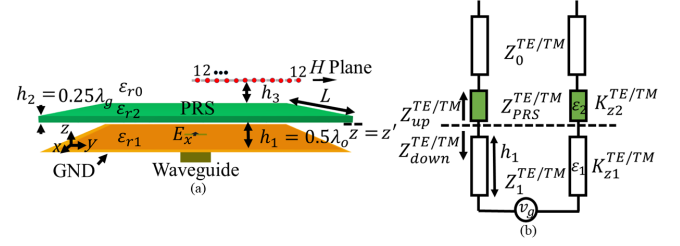


Fig. 1. (a) Configuration of the conventional FPCA; The antenna has a square geometry with an area of $(L \times L = 5.32\lambda_0 \times 5.32\lambda_0)$; λ_0 is the wavelength at the resonant frequency of 13.3 GHz. (b) TEN model of the structure in Fig. 1(a).

ing the metasurface unit cells capable of providing quasi-linear phase compensations. Then, we design the proposed PCMS and place it in the near-field region of the conventional FPCA to construct the proposed antenna. The simulation and measurement results showcase that the proposed antenna achieves a wider 3 dB gain bandwidth of approximately 35% with a peak gain of 20.1 dBi. To the author's knowledge, the achieved bandwidth of 35% is the widest bandwidth reported for a class of FPCA designed with planar PCMSs.

Section II of the paper studies the frequency-dependence of complex modes in the FPCA cavity. Section III presents a PCMS which helps the FPCA achieves phase uniformity over its aperture over a wide operation bandwidth. Section IV validates the proposed methodology through simulations and measurement results. Finally, Section V concludes the paper.

II. DISPERSION ANALYSIS OF THE CONVENTIONAL FPCA

We begin by analyzing the conventional FPCA structure with a dielectric PRS to study the dispersion behavior of leaky wave modes. Fig. 1(a) illustrates the configuration of an FPCA, which is designed to work at the resonant frequency of 13.3 GHz. We employ the dielectric PRS of thickness $\lambda_g/4$ (where λ_g is a guided wavelength in the dielectric material, given by $\lambda_g = \frac{\lambda_0}{\sqrt{\epsilon_r}}$, with a metallic ground plane positioned $\lambda_0/2$ away from the PRS. The medium between the PRS and the ground and above the PRS is air ($\epsilon_{r1} = 1.006$). We select this substrate thickness and the separation distance based on [3], [33], [37], [44], [45] to construct the resonant cavity at 13.3 GHz, achieving maximum FPCA directivity [46]. We employ a WR-62 waveguide-fed slot antenna at the center of the ground plane to feed the FPCA. The waveguide's operating mode is TE_{10} , which excites the E_x , H_y , and H_z fields. The source excites a pair of nearly degenerate TE/TM leaky-wave modes, which propagate inside the cavity with complex transverse propagation constant $k_t = \beta - j\alpha$, where β and α represents the phase and attenuation constants (leakage rate), respectively, and $k_t = \sqrt{k_x^2 + k_y^2}$. In the following, we apply a simple but accurate approach based on a transverse equivalent network (TEN) [47]–[49] to examine the behavior of the fundamental TE/TM leaky-wave modes inside the cavity.

A. TEN Analysis of the Conventional FPCA

We mostly employ the method described in [47]–[51] to implement TEN analysis on the FPCA structure. Assuming

that the FPCA has an infinite PRS and ground plane along the transverse plane, the equivalent transmission line model of the FPCA is depicted in Fig. 1(b). We model the PRS, the medium between the PRS and the ground plane, and the medium above the PRS with equivalent TE/TM transmission lines with characteristic impedances of Z_{PRS} , Z_1 and Z_0 . The short circuit represents the metallic ground of the FPCA, respectively. The equivalent TE/TM transmission lines' dispersion relation can be used to find the leaky wave's propagation constant as a function of frequency. We apply the transverse resonance condition at $z = z'$:

$$Z_{\text{up}} + Z_{\text{down}} = 0 \quad (1)$$

where Z_{up} and Z_{down} represent the equivalent impedance looking up and down, can be calculated from the transmission line theory as follows:

$$Z_{\text{up}} = Z_{\text{PRS}} \frac{Z_0 + jZ_{\text{PRS}} \tan(k_{z2}h_2)}{Z_{\text{PRS}} + jZ_0 \tan(k_{z2}h_2)} \quad (2)$$

$$Z_{\text{down}} = jZ_1 \tan(k_{z1}h_1) \quad (3)$$

where Z_0 , Z_1 , and Z_{PRS} represent the characteristics impedances of the TE/TM waves in their respective media ($Z_i^{\text{TE}} = \frac{\eta_i k_i}{k_{zi}}$ and $Z_i^{\text{TM}} = \frac{\eta_i k_{zi}}{k_i}$) [47], and the k_{zi} is the longitudinal propagation constant given by $k_{zi} = \sqrt{\epsilon_{ri}k_0^2 - k_t^2}$ for medium $i = 0, 1, 2$. Applying numerical methods to solve (1) using (2) and (3) provides the dispersion of the transverse (k_{t1}) and longitudinal (k_{z1}) complex propagation constants inside the cavity. However, a more convenient analytical expression can be obtained by linearly approximating the $\tan(x)$ function in eq. (3) around its zeros in the neighborhood of $x \pm \pi$ [49]. This simplifies (1) to:

$$k_{z1} = j \frac{Z_{\text{up}}}{h_1 Z_1} + \frac{\pi}{h_1} \quad (4)$$

Approximating the cavity load impedance (Z_{up}) value around broadside ($Z_{\text{up}} = \eta_0/\epsilon_{r2}$) – assuming $h_2 = \lambda_g/4$ in (2) – and substituting Z_1 for the TE and TM cases into Eq. (4), we find the propagation constants for the TE and TM modes within the cavity [49]:

$$k_{z1}^{\text{TE}} = \frac{\epsilon_{r1}k_0}{\pi\epsilon_{r2} \left(\frac{2h_1\sqrt{\epsilon_{r1}}}{\lambda_0} \right)^2} \frac{1 + \frac{2\pi h_1\epsilon_{r2}}{\lambda_0} + j}{1 + \left(\frac{2\pi h_1\epsilon_{r2}}{\lambda_0} \right)^2}, \quad (5)$$

and

$$k_{z1}^{\text{TM}} = \frac{\sqrt{\epsilon_{r1}}k_0}{\left(\frac{4h_1\sqrt{\epsilon_{r1}}}{\lambda_0} \right)} \left(1 \pm \sqrt{1 + 8j \frac{\epsilon_{r1}h_1}{\lambda_0 \pi\epsilon_{r2}}} \right) \quad (6)$$

Moreover, the cavity supports the propagation of an undesired spurious TM_0 or quasi-TEM leaky-wave mode, which is similar to the TEM mode of the parallel plate waveguide (PPW) [52]. The dispersion of the TM_0 mode can be calculated by [53]

$$k_{z1}^{\text{TM}} = \sqrt[3]{jk_0^2 \frac{\epsilon_{r2} - 1}{\epsilon_{r2}^2 h_0}} \quad (7)$$

We calculate the associated transverse propagation constants for the quasi-TEM, TE, and TM leaky modes using $k_t =$

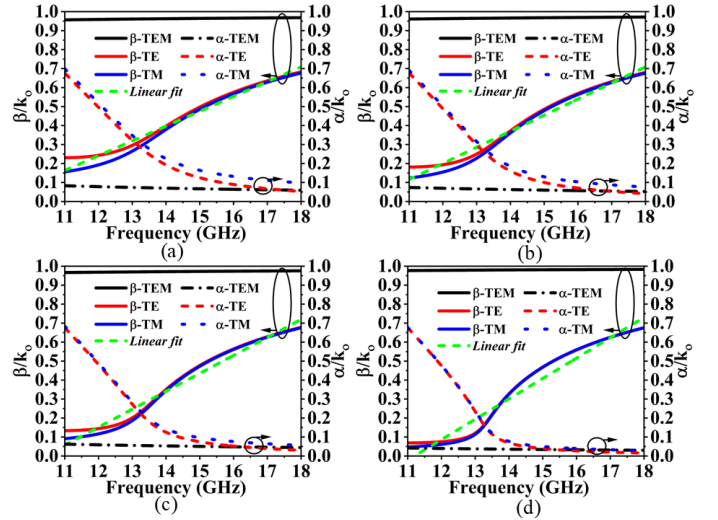


Fig. 2. Normalized complex wave numbers for leaky waves propagating transversally. Propagation constant β (solid lines) and attenuation constant α (dash-dotted lines) are shown for TEM (black), TE (red), and TM (blue) modes with dielectric permittivities: (a) $\epsilon_r = 3.5$, (b) $\epsilon_r = 4.5$, (c) $\epsilon_r = 6.15$, (d) $\epsilon_r = 10.2$. Green dash lines show linear fits $\beta = mf + c$ to TE/TM data. The linear dispersion parameters (slope, y-intercept) are $m = [0.0847, 0.0779, 0.0945, 0.107]$ and $c = [-0.815, -0.693, -0.982, -1.201]$ for panels (a)-(d) respectively.

$\sqrt{\epsilon_{r1}k_0^2 - k_{z1}^2}$. To analyze the dispersion characteristics of leaky wave modes inside the cavity of the FPCA, we consider a dielectric PRS with varying relative permittivities (ϵ_{r2}) of 3.5, 4.5, 6.15, and 10.2. The dispersion relationships are plotted in Fig. 2 to illustrate the frequency-dependent behavior of these modes in the Ku-band. We first make observations on the TEM mode. We observe that throughout the observed frequency range, the normalized phase constant (β/k_0)— k_0 is the free space wavenumber which varies with frequency—of the TEM mode exceeds 0.95, while its attenuation constant (leakage rate, α/k_0) remains extremely small. This indicates the TEM wave propagates with a phase velocity close to the speed of light in the cavity, experiencing minimal dispersion and negligible leakage, allowing the TEM wave to travel a relatively long distance inside the cavity. The sustained propagation of the TEM mode results in a large effective radiating aperture. We now direct our attention to the 1st-order TE and TM modes. We can see that the TE and TM leaky-wave modes exhibit nearly identical dispersion curves, indicating comparable propagation and attenuation constants within the cavity for throughout the observed frequency range. These modes display higher attenuation constants below the cutoff frequency of 13.3 GHz, causing them to radiate quickly into the central region above the feed. This phenomenon reduces the transverse propagation of waves and shrinks the effective radiating aperture. Above the cutoff frequency, however, the attenuation constants decrease significantly and approach the attenuation constants of the TEM modes, allowing the modes to propagate farther transversely in the cavity with less radiative leakage. This behavior suggests that, for these two modes, operating above the cutoff frequency is critical for maximizing the aperture size and radiation efficiency. Also, above the cutoff frequency, the TE and TM waves propagate with a

quasi-linearly increasing normalized phase constant (β/k_0) against frequency. This relationship holds true for different permittivity values. Using linear regression, we model the normalized β as a linearly increasing function of frequency (f) with the best-fit equation $\beta = mf + c$ where m quantifies the rate of increase and c is the intercept. The results demonstrate that for all permittivity values, the normalized β of both TE and TM waves closely approximates the fitted equation's β . The results clearly indicate that the $\frac{\Delta\beta}{\Delta f} \approx m_0$ from 13 GHz to 18 GHz. This observation holds significant importance, as it reveals that the aperture phase of the dominant electric field in the TE and TM leaky waves undergoes a nearly constant rate of change against frequency $\frac{\delta\phi(r)}{\delta f} \approx m_1(r)$, where r represents the radial position. We will elaborate on this in the next section.

B. HFSS Simulations and Extraction of Complex Propagation Constants

In the previous subsection, we analyzed the frequency-dependent behavior of the calculated leaky-wave wavenumbers. Here, we examine how this wavenumber behavior manifests within the FPCA aperture field. First, we verify whether the theoretical wavenumber values from the last section align with those obtained from full-wave simulation. To accomplish this, we simulate the FPCA in HFSS and observe the dominant field components across transverse planes. We then fit the simulated field profiles in each plane using a non-linear least squares optimization to obtain estimates on the numerical wavenumber. The results demonstrate that these numerically extracted values show good agreement with the theoretically calculated wavenumbers from the preceding section. Subsequently, we compute the FPCA aperture phase using the numerically calculated wavenumbers, and demonstrate its quasi-linear variation with frequency.

We simulate the FPCA shown in Fig. 1(a) using a PRS with a dielectric constant ϵ_r of 4.5. According to the leaky-wave interpretation of FPCAs [54], the cavity supports the fundamental quasi-TEM^x mode along with higher-order TM^x and TE^y modes, where the superscripts indicate the propagation direction. Recent works [42], [43], [55] relate these modes to their characteristic field distributions and propagation behavior. Our simulated near-field results show that E_x , E_z , H_y , and H_z are the non-zero field components both inside and outside the cavity, for both resonant and non-resonant frequencies. Specifically, in the E-plane, the field distributions exhibit characteristics (E_x , E_z , H_y) associated with TM^x propagation, while in the H-plane, the field distributions exhibit characteristics (E_x , H_y , H_z) associated with TE^y propagation [42]. These observations are consistent with the modal interpretation of the FPCA cavity involving the fundamental quasi-TEM^x mode and the first TM^x and TE^y leaky-wave pairs.

To examine the behavior of the dominant field components within the FPCA cavity and to relate it to the analysis presented in Fig. 2, we plot the magnitude and phase of these field components outside the cavity, at a distance of $0.25\lambda_0$ above the PRS for three different frequencies, as shown in Fig. 3. The figure provides a top view of the antenna,

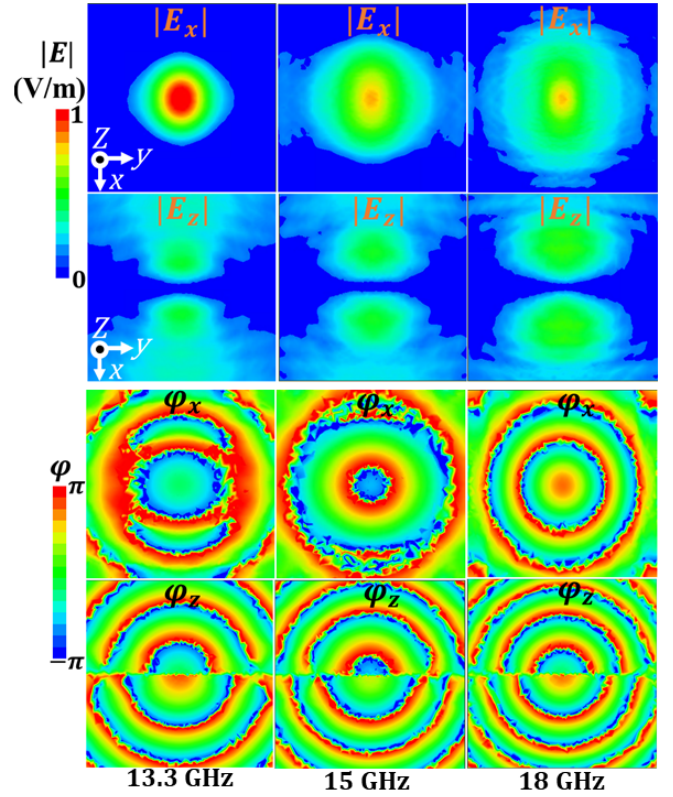


Fig. 3. 2D complex electric field distributions of the dominant E_x and E_z components of the (TE and TM) and TEM leaky-wave modes, respectively, plotted above the PRS at the vertical distance of $z = 0.25\lambda_0$.

showing the spatial variations of $|E_x|$, $|E_z|$, ϕ_x , and ϕ_z . Fig. 3 (first row) shows the behavior of the $|E_x|$ component, which is dominant in the TE/TM modes. It reveals rapid energy leakage from the cavity at the resonant frequency (13.3 GHz), a consequence of the higher attenuation constants of these modes as indicated in Fig. 2. The paired leaky waves propagate symmetrically in the E- and H-planes, yielding a near-circular field distribution. Moreover, at the nonresonant frequencies (15 GHz and 18 GHz), the energy is more spread out on the antenna aperture due to the TE and TM modes' smaller attenuation constants (leakage rate). Conversely, $|E_z|$ (second row) exhibits propagation along the $\pm x$ directions on the E-plane, consistently observed at the resonant and nonresonant frequencies. This direction aligns with the TEM mode's minimal attenuation constant (as discussed in the last section), enabling sustained energy confinement within the cavity's E-plane over a large area. Moreover, Fig. 3 (third row) reveals the wrapped phase, ϕ_x , which exhibits a nearly circularly symmetric distribution near the resonant frequency and becomes even more circularly symmetric away from resonance. As frequency increases, the aperture phase varies more rapidly, which is evident from the growing number of annular rings in the phase profile. Similarly, ϕ_z also exhibits phase variations at both resonant and nonresonant frequencies, as demonstrated in the bottom row of Fig. 3. In the following, we probe the simulated complex electric field in the transverse planes and apply a non-linear least squares optimization to determine the numerical wavenumber values. For brevity, we

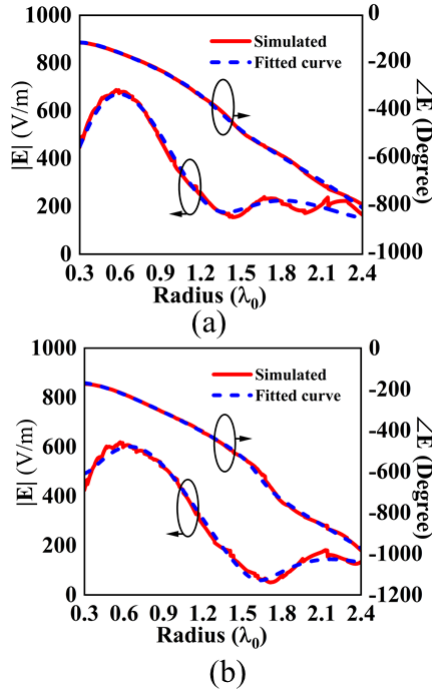


Fig. 4. Comparison of numerically computed and simulated complex electric near-field at the E-plane inside the cavity at (a) 13.3 GHz and (b) 15 GHz. (λ_0 is the wavelength at the resonant frequency.)

probe the electric field inside the cavity along the E-plane and outside the cavity along the H-plane.

Considering the field symmetry in the E-plane, we probe the field within the region inside the cavity defined by $0 < x < L/2$. We extract the total near-electric field data on a straight line inside the cavity just below the PRS (at $z = 0.49\lambda_0$). To extract a more uniform field from the simulations, we probe the fields in the region $(0.3\lambda_0 - 2.4\lambda_0)$, avoiding the feeding source and unwanted reflections from the edges. We then use a nonlinear least-squares optimization to numerically fit the extracted electric field to the waveform. Since in the E-plane the TEM^x wave and one of the leaky waves TM^x propagate inside the cavity, we suppose two traveling waves propagating inside the cavity with unknown complex wavenumbers of k_{t1} and k_{t2} , respectively. We have:

$$E_{\text{total}}(x) = A_1 e^{-k_{t1}(x)} + A_2 e^{-k_{t2}(x)}, \quad (8)$$

where A_1 and A_2 are the complex amplitudes of the TEM^x and TM^x traveling waves respectively, and $E_{\text{total}}(x)$ represents the numerically computed complex electric field. The initial lower and upper bounds of these unknowns can be estimated by simulated near-field data. The fitting process gives the values of four unknowns (A_1 , k_{t1} , A_2 , and k_{t2}) that best fit the amplitude and phase of the simulated and numerical electric field simultaneously. We summarize the fitted and theoretical dispersion characteristics of the modes in Table I. For completeness, the table includes the dispersion characteristics of the TE cavity wave associated with Fig. 5.

Fig. 4 compares the simulated amplitude and phase of the electric field and their corresponding fitted curves at the

TABLE I
COMPARISON OF THE THEORETICAL AND FITTED DISPERSION CHARACTERISTICS OF THE MODES: PROPAGATION CONSTANT (β/k_0) AND ATTENUATION CONSTANT (α/k_0)

Freq. (GHz)	Mode	β/k_0		α/k_0	
		Theory	Fit	Theory	Fit
13.3	TEM^x	0.965	1.04	0.064	0.079
13.3	TM^x	0.265	0.30	0.264	0.245
13.3	TE^y	0.281	0.269	0.245	0.285
15.0	TEM^x	0.968	1.08	0.056	0.078
15.0	TM^x	0.486	0.473	0.215	0.223
15.0	TE^y	0.489	0.525	0.156	0.147

resonant frequency (13.3 GHz) and one of the non-resonant frequencies (15 GHz) for brevity. It shows that the sum of the two traveling waves fits the near-field electric field amplitude and phase very well. In Fig. 4(a), the numerically calculated values of normalized attenuation and phase constants ($\alpha_1 = 0.079k_0$, $\beta_1 = 1.04k_0$) and ($\alpha_2 = 0.308k_0$, $\beta_2 = 0.245k_0$) approximately match the theoretically calculated ($\alpha_{\text{TEM}} = 0.0648k_0$, $\beta_{\text{TEM}} = 0.9654k_0$) and ($\alpha_{\text{TM}} = 0.2648k_0$, $\beta_{\text{TM}} = 0.2652k_0$) values of the TEM and TM waves respectively. Also, the magnitudes of the first and second traveling waves are $|A_{\text{TEM}}| = 52.25$ and $|A_{\text{TM}}| = 1151$. Similarly, in the case of Fig. 4(b), the numerically calculated values of normalized attenuation and phase constants ($\alpha_1 = 0.078k_0$, $\beta_1 = 1.08k_0$) and ($\alpha_2 = 0.2230k_0$, $\beta_2 = 0.473k_0$) approximately match the theoretically calculated ($\alpha_{\text{TEM}} = 0.0561k_0$, $\beta_{\text{TEM}} = 0.9680k_0$) and ($\alpha_{\text{TM}} = 0.2150k_0$, $\beta_{\text{TM}} = 0.4860k_0$) values of the TEM and TM waves respectively. The magnitudes of the first and second traveling waves, in this case, are $|A_{\text{TEM}}| = 379$ and $|A_{\text{TM}}| = 1143$. Thus, this demonstrates that the wavenumber values obtained from the TEN model approximately match the wavenumbers extracted from the simulated field profile. Furthermore, it confirms that the TM leaky wave remains the dominant mode across all frequencies in the E-plane. Furthermore, we probe the simulated complex electric field in the H-plane outside the cavity at a height of $h_3 \approx 0.25\lambda_0$ above the PRS, within the region $0.3\lambda_0 < y < 2.4\lambda_0$. In this plane, the TE mode propagates with a dominant E_x component. The theoretical profile of the TE mode, namely

$$|E_x|(y) = A_3 e^{-\alpha_{\text{TE}}(y)} \quad (9)$$

$$\angle E_x(y) = -\beta_{z,\text{TE}}h_1 + \theta_{\text{PRS}} - \beta_{z,\text{TE}}h_3 - \beta_{t,\text{TE}}y + \theta_0 \quad (10)$$

where $\beta_{z,\text{TE}}$ and $\beta_{t,\text{TE}}$ represent the longitudinal and transverse phase constants of the TE cavity wave, respectively; h_1 and h_3 denote the vertical cavity height inside and outside the cavity, respectively; θ_{PRS} is the transmission phase of the PRS; y represents the radial propagation; and θ_0 is the phase constant-offset parameter obtained by fitting the analytical expression to the simulated aperture phase distribution. We calculate θ_{PRS} using $\theta_{\text{PRS}} \approx -k_0\sqrt{\varepsilon_r}t$, where k_0 is the free-space wavenumber, ε_r is the dielectric constant, and t is the slab thickness. We have verified that over the operating band, this approximation provides sufficiently accurate phase shifts

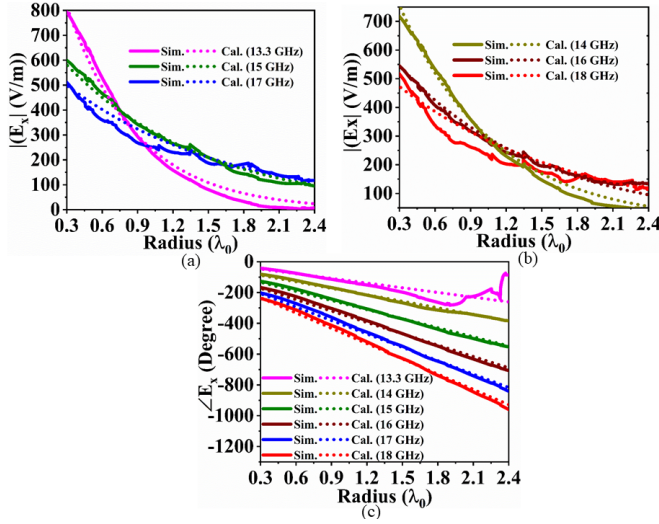


Fig. 5. Comparison of numerically computed and simulated complex electric near-field at the H-plane outside the cavity: (a-b) Amplitude comparison and (c) Phase comparison. (λ_0 is the wavelength at the resonant frequency.)

with a deviation of less than 3° . A_3 is a constant amplitude coefficient obtained from fitting the aperture-field magnitude distribution.

Fig. 5 compares the simulated and calculated magnitudes and phases of the E_x in the H-plane. It can be seen that the calculated magnitudes and phases approximately agree with the simulation results, validating the effectiveness of the TEN model in determining the complex wave numbers and, eventually, the antenna aperture's magnitude and phase. It is important to point out here that the aperture phase and magnitude of the FPCA have been directly extracted from HFSS simulations in several previous studies [32], [34]. In this context, the proposed method based on calculation enables the rapid determination of the antenna's near-fields, provided that the FPCA design parameters such as cavity height, PRS permittivity are known. Moreover, the HFSS simulations show that the phase slightly deviates from the calculated results at the resonant frequency for the region $r > 1.5\lambda_0$. This is perhaps due to the multiple reflections inside the resonant cavity, though it does not significantly affect the antenna performance, as this region has less energy than the central region of the antenna due to higher attenuation constants of TE leaky wave. Furthermore, the unwrapped phases in Fig. 5(c) are transversely nonuniform, and their dispersion increases with respect to frequency. For example, at the resonant frequency of 13.3 GHz, the aperture phase varies from -28° to -267° , which varies more from -177° to -837° at the nonresonant frequency of 17 GHz. Furthermore, the phase variation with respect to frequency ($\partial \angle E_x / \partial f$) is minimal near the aperture's center and increases significantly as waves propagate transversely toward the edge.

To effectively show the FPCA's aperture phase variation with respect to frequency, we plot the phase in the H-plane (the yoz-plane) at 12 different spatial locations with a separation distance of approximately $0.18\lambda_0$. The region spanning ($0 < y < 2.1\lambda_0$) accounts for nearly 95% of the

TABLE II
VIRTUALLY PROBED PHASES OF E_x ACROSS THE H-PLANE OUTSIDE THE CAVITY FROM 13 GHz TO 18 GHz

Spatial Location no	Electrical Distance	θ_{13} (deg)	θ_{14} (deg)	θ_{15} (deg)	θ_{16} (deg)	θ_{17} (deg)	θ_{18} (deg)
1	$0.08\lambda_0$	-12	-67	-109	-147	-179	-214
2	$0.26\lambda_0$	-20	-77	-122	-162	-197	-232
3	$0.44\lambda_0$	-37	-98	-148	-192	-231	-270
4	$0.62\lambda_0$	-58	-124	-181	-232	-277	-319
5	$0.80\lambda_0$	-82	-152	-219	-278	-328	-380
6	$0.98\lambda_0$	-102	-179	-257	-321	-385	-436
7	$1.15\lambda_0$	-119	-209	-296	-370	-444	-503
8	$1.33\lambda_0$	-144	-238	-334	-417	-496	-567
9	$1.50\lambda_0$	-184	-268	-373	-469	-552	-631
10	$1.69\lambda_0$	-232	-300	-413	-519	-614	-695
11	$1.86\lambda_0$	-280	-321	-453	-565	-662	-758
12	$2.04\lambda_0$	-311	-335	-487	-616	-722	-824

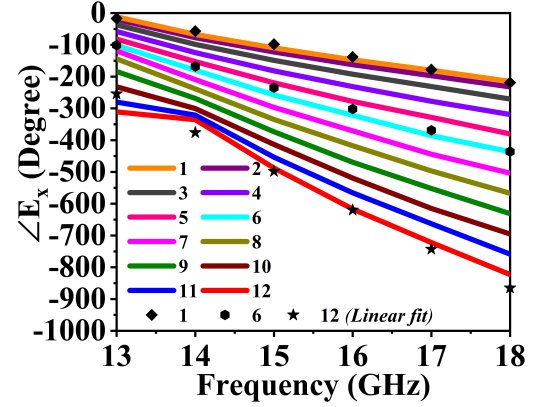


Fig. 6. Unwrapped phase variations of the E_x across twelve spatial locations in the H-plane as a function of frequency. The black symbols (e.g., circles and squares) show linear fits to Locations 1, 6, and 12 with the following parameters (slope m , intercept c): Location 1: $(-40.4, 508.3)$; Location 6: $(-66.9, 768.9)$; Location 12: $(-122.4, 1338)$.

normalized power $|E_x|^2$, significantly lower near the antenna edges. Table II summarizes the probed 1D phases, which are plotted in Fig. 6. Note that in Table II, the numbers in the subscript of θ (13, 14, ..., 18) are the frequencies in GHz where the phase was taken. This is only for Table II and is not related to the subscripts used in Eq. (11). Fig. 6 shows that the phases quasi-linearly correlate with the frequencies, which becomes more pronounced for the region of 14 GHz to 18 GHz after the resonant frequency of 13.3 GHz. For example, the probed phases at the spatial Location 1 change from -12° to -214° when the frequency changes from 13 GHz to 18 GHz. The figure also shows linear fits to the data at Locations 1, 6, and 12 (marked with black symbols), using the equations given in the caption. Comparing this phase variation with the ideal linear phase response (obtained from the equation of a straight line), the resulting phase error between these two curves is merely 10° . In a purely linear system, such phase difference would be

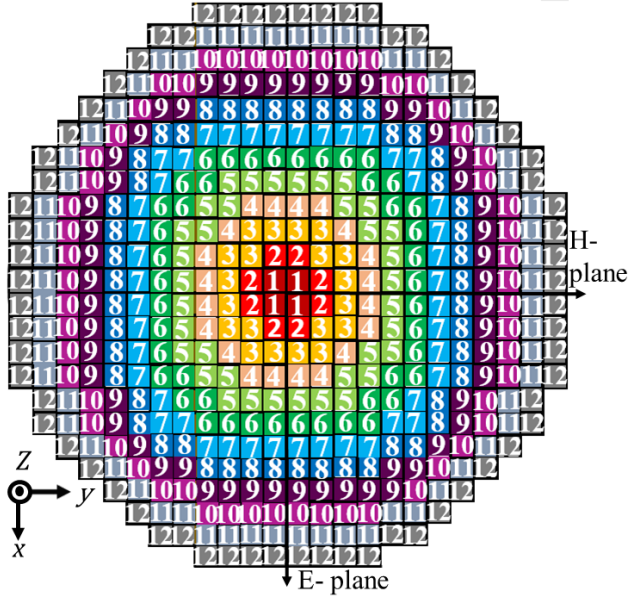


Fig. 7. 2D representation of the virtually probed phases of E_x outside the cavity.

constant, producing constant phase delays, which at location 1 tend to stabilize somewhat at the higher frequencies within a phase error of 10° , suggesting a quasi-linear relationship overall. This phase error remains approximately 30° for the first 11 locations and reaches approximately 50° for Location 12 for the frequency region of 14 GHz to 18 GHz when comparing with the equations of straight lines in each case. This behavior occurs because the FPCA aperture's quasi-linear phase variation against frequency stems from the quasi-linear behavior of phase constants β_{TE} as shown in Fig. 2. In the next section, we design a novel wideband PCMS capable of correcting these phase non-uniformities at the resonant and non-resonant frequencies for the dominant electric field component (E_x) of TE and TM leaky waves.

III. DESIGN OF A PHASE CORRECTING METASURFACE FOR THE FPCA

In this section, we design a PCMS that functions as a transparent structure, providing minimal reflections while maintaining precise dispersive phase control across its aperture. To achieve accurate phase compensation for the dominant TE and TM leaky waves across both resonant and nonresonant frequencies, we implement the PCMS in the near-field of the FPCA at a vertical height of $h_3 = 0.25\lambda_0$ above the PRS, as shown in Fig. 8 (where λ_0 is the free-space wavelength at the center frequency of 13.3 GHz). The phase profile requiring correction, derived from the nonuniform phase of E_x at h_3 presented in the previous section, is rotationally symmetric (see Figs. 3 and 7). Consequently, the required PCMS is a 2D structure populated by axially placed unit cells. The additional phase shift each unit cell must provide is calculated as:

$$\Delta\theta_n(\omega) = \theta_{12}(\omega) - \theta_n(\omega) + \theta_0(\omega), \quad n = 1, 2, \dots, 12 \quad (11)$$

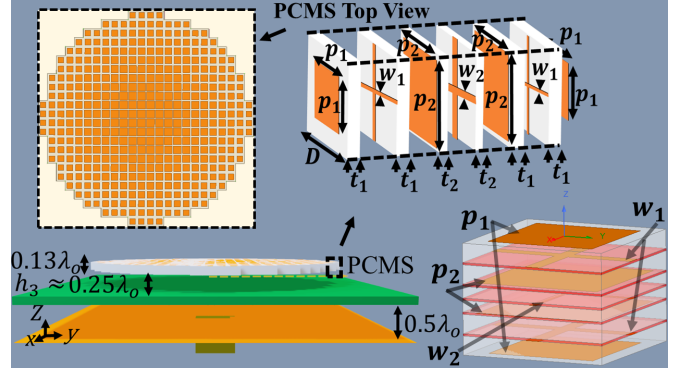


Fig. 8. Configuration of the proposed FPCA employing the PCMS in its near field; the inset above the proposed antenna shows the top view of the PCMS; the top right inset shows the PCMS unit cell where metallic patterns are etched on six ultrathin dielectric boards of thicknesses given as: $t_1 = 0.508$ mm; $t_2 = 0.305$ mm; The bottom right shows a cross-sectional view of the PCMS unit cell, where the red sheets are the bonding films made of RO4450F (with a relative permittivity $\epsilon_r = 3.52$) and a thickness of 0.065 mm.

Here, $\Delta\theta_n$ is the frequency-dependent phase shift to be implemented at the 12 discrete spatial locations (marked in Fig. 1(a)), calculated relative to the phase θ_{12} at Location 12 (the point of greatest phase delay). The quantities $\theta_n(\omega)$ and $\theta_{12}(\omega)$ are extracted directly from the aperture phase distributions shown in Fig. 6. The frequency-dependent reference phase constant $\theta_0(\omega)$ is introduced to facilitate the practical realization of these phase delays by the PCMS unit cells. The selected values of $\theta_0(\omega)$ are -30° , -5° , 40° , 60° , 40° , and 40° at 13.3, 14, 15, 16, 17, and 18 GHz, respectively. The proposed integrated antenna is shown in Fig. 8. Its PCMS aperture consists of unit cells designed to implement the spatially and frequency-dependent phase shifts, $\Delta\theta_n(\omega)$. After interacting with this tailored PCMS aperture, the cavity waves achieve a nearly uniform phase distribution.

A. Unit Cell Design

The essential constituent of the PCMS is the phase-compensating unit cell, whose operational characteristics govern the overall functionality of the PCMS. Knowing the FPCA aperture's quasi-linear phase variations with respect to frequency and radial locations, specialized unit cells exhibiting linearly varying phases as a function of frequency are the best candidates for the FPCA's aperture phase corrections across wideband. This is a critical aspect of designing a wideband PCMS, which has not been considered in the earlier reported research [32], [36], [38], which mainly focus on achieving the desired compensation at the resonance frequency of the FPCA. In this work, we design the PCMS unit cells to provide the required linearly varying phase shifts with good transmission magnitude over a wide frequency range.

The inset of Fig. 8 (in the top right) shows the physical structure of the proposed PCMS unit cell, where the dielectric layers are spaced out to show the metallic patterns. The unit cell structure comprises six ultrathin dielectric substrates made of Rogers RO4003C ($\epsilon_r = 3.4$; $\tan\delta = 0.0027$) and seven metallic layers of nonresonant capacitive patches and inductive grids. The topmost and bottommost square patches exhibit

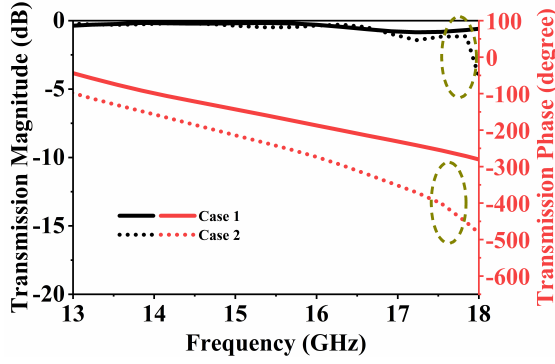


Fig. 9. The transmission coefficients of the PCMS unit cells when designed for the two different slopes. The ovals indicate the region near the cutoff frequency in Case 2 where the transmission magnitude drops sharply and the phase changes rapidly.

identical dimensions with lengths of p_1 , as shown in Fig. 8. The middle square patches are similar, with lengths of p_2 . Similarly, the inductive grids located at the topmost and bottommost positions possess identical dimensions, with grid widths of w_1 . In contrast, the inductive grid situated in the middle features a distinct width, referred to as w_2 . The overall thickness of the unit cell is $0.13\lambda_0$, and its lateral dimension is $D = 0.18\lambda_0$ which is similar to the probing distance between the two consecutive spatial locations given in the last section. The proposed PCMS unit cell provides the fourth-order ($\frac{N+1}{2}$) bandpass filter characteristics with linearly varying phases with respect to frequency, where N represents the number of metallic layers. We adopt the fourth-order filter topology because it provides linearly varying responses with higher phase compensation in a wideband than the lower-order filters. We design the proposed unit cell following the generalized method provided in [56] to synthesize the desired filtering response. Ref. [56] presents a comprehensive analytical design and synthesis procedure to realize linearly varying phase shifts in a metasurface unit cell. Given a target center frequency (f_0), fractional bandwidth (BW), filter order, and desired response type, this procedure derives the metasurface unit cell's lumped component values and geometrical parameters to realize any filter response. We employ this synthesis procedure to realize PCMS's unit cells with high transmission efficiency and linearly varying phase shifts across our target frequency band, aiming to transform the FPCA's nonuniform aperture phase to uniform.

We first look at the ability of the proposed unit cell to provide linearly varying phases against frequency (or constant group delays ($\tau_g = \frac{d\phi(f)}{df}$) over the bandwidth of interest, which can be further detuned to match the required quasi-linear $\Delta\theta_n$. Fig. 9 shows two examples (Case 1 and Case 2) for which we implement a linear phase filter following the analytical procedure of [56]. The metallic layers dimensions are ($p_1 = 3.25$ mm; $p_2 = 3.37$ mm; $w_1 = 0.6$ mm and $w_2 = 0.8$ mm) for Case 1 and ($p_1 = 3.55$ mm; $p_2 = 3.65$ mm; $w_1 = 0.9$ mm and $w_2 = 0.9$ mm) for Case 2. We implement the steeper phase slope against frequency in Case 2 than in Case 1 to observe its impact on the filter's bandwidth. The steeper slope, which corresponds to a larger group delay

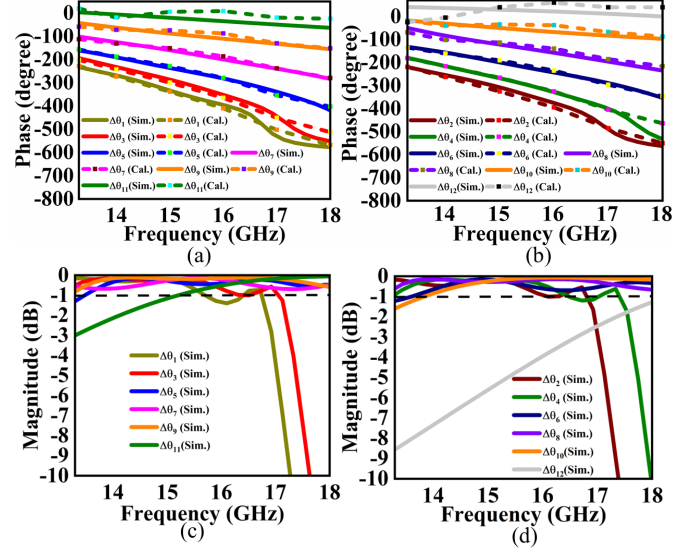


Fig. 10. Comparison of the simulated and the required transmission phases and magnitudes at the 12 spatial locations.

($\tau_{g2} = 160$ ps vs. $\tau_{g1} = 120$ ps), reduces the filter's relative bandwidth by bringing the upper cutoff frequency near 18 GHz [57]. Consequently, while both implementations provide a linearly varying phase with relatively constant group delay, the design with the steeper slope shrinks the operating bandwidth and causes a sharp attenuation in transmission magnitude near this cutoff. Therefore, designing linear phase filters for steeper slopes degrades the transmission bandwidth at higher frequencies.

Fig. 10 shows the transmission coefficients of the proposed unit cells aiming to provide the desired $\Delta\theta_n$ at the resonant (13.3 GHz) and nonresonant frequencies (14 GHz to 18 GHz). The desired $\Delta\theta_n$ is calculated from (11). Table III lists the dimensions of the metallic layers of unit cells achieving the desired responses. Fig. 10(a-b) show the transmission phases of designed unit cells, which are compared against the calculated $\Delta\theta_n$. Fig. 10(c-d) provide the corresponding transmission magnitudes of these unit cells. It is evident from Fig. 10 that the unit cells mostly possess the desired $\Delta\theta_n$ with good transmission magnitudes ($|S_{21}| > -1$ dB) with the following exceptions. Unit Cells 1-4 show a deviation of 50° from the ideal required phases with attenuating transmission magnitude after 17 GHz; Unit Cells 11-12 have relatively low transmission magnitude near the resonant frequency, which increases moving towards the nonresonant frequencies. The Unit Cells 1-4 have the upper cutoff frequency near 17 GHz, where the phases abruptly deviate from the ideal linear phase and transmission magnitudes undergo attenuations. This can be further improved by designing a higher-order filter, requiring more metallic layers and increasing the computational complexity. The Unit Cells 11-12 are located in the outer region of the PCMS, where the energy naturally weakens. Thus, these unit cells do not significantly impact the overall performance of the FPCA.

TABLE III
GEOMETRIC CHARACTERIZATION OF METALLIC PATTERN DIMENSIONS
IN UNIT CELL DESIGN

Spatial Location (Ring no)	P_1 (mm)	P_2 (mm)	W_1 (mm)	W_2 (mm)
1	3.7	3.76	0.65	0.1
2	3.62	3.76	0.55	0.1
3	3.58	3.72	0.65	0.1
4	3.55	3.68	0.65	0.1
5	3.4	3.58	0.45	0.1
6	3.3	3.5	0.45	0.45
7	3.21	3.4	0.45	0.5
8	3.1	3.2	0.5	0.6
9	3	3.02	0.4	0.5
10	1.8	1.8	0.05	0.3
11	0.4	0.4	0.05	0.2
12	0.3	0.5	0.2	0.2

IV. SIMULATION AND EXPERIMENTAL RESULTS

We simulate the proposed FPCA in HFSS to examine the impact of PCMS-generated additional phase shifts on the antenna's nonuniform aperture phase. The full-structure simulation requires approximately 96 hours and 300 GB of memory. Convergence is achieved by setting the maximum Delta S to 0.02 and requiring two consecutive converged passes. With the additional phase shifts, the outgoing waves leaving the PCMS would ideally propagate with a uniform aperture phase. Fig. 11 plots the electric field phase on the H-plane, along a line placed $0.1\lambda_0$ above the PCMS. The phase profiles for the FPCA with and without the PCMS are compared. It is observable that the inclusion of PCMS above the FPCA remarkably improves the nonuniformity of the aperture phase of the FPCA. The uniform phase region, defined as a region with a phase difference of less than 50° [35], at different frequencies for FPCA with PCMS is: $(-1.2\lambda_0 < y < 1.2\lambda_0)$ for 13.3 GHz, $(-1.6\lambda_0 < y < 1.6\lambda_0)$ for 14 GHz, $(-1.9\lambda_0 < y < 1.9\lambda_0)$ for 15 GHz, $(-1.7\lambda_0 < y < 1.7\lambda_0)$ for 16 GHz, and $(-1.7\lambda_0 < y < 1.7\lambda_0)$ for 17 GHz. The maximum aperture phase variation occurs near the edges of the structure, which at 13.3 GHz, 14 GHz, 15 GHz, 16 GHz, and 17 GHz are 100° , 75° , 80° , 96° and 88° respectively, are significantly improved from 200° , 270° , 375° , 444° , 513° without the PCMS. In Fig. 12, we plot the $\angle E_x$ to visualize the phase variation of the radiated waves at 13.3 GHz, 15 GHz, and 17 GHz. It can be seen that the aperture phase is considerably more uniform in the presence of the PCMS at all frequencies. This ensures highly directive radiation patterns into far-field.

We fabricate a prototype of the proposed antenna comprising the PCMS, PRS, and metallic ground with a 3D printed support, as separately shown in Fig. 13(a). We stack the dielectric layers of the PCMS with the RO4450F bonding films, where each bonding film has a thickness of 0.065 mm. We assemble the proposed antenna by placing the PRS at

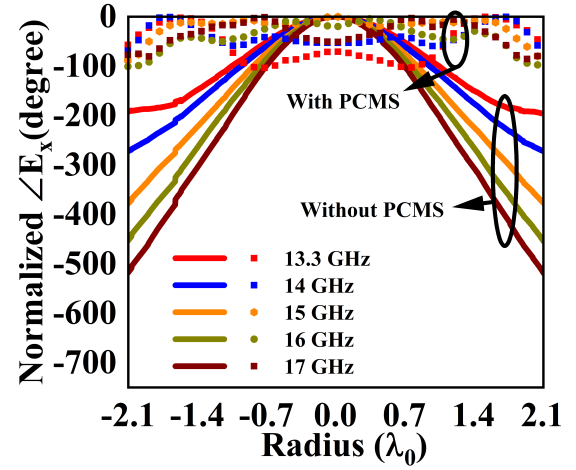


Fig. 11. 1-D near-field phase distribution comparison for the FPCA with and without the PCMS; The λ_0 is the wavelength at 13.3 GHz.

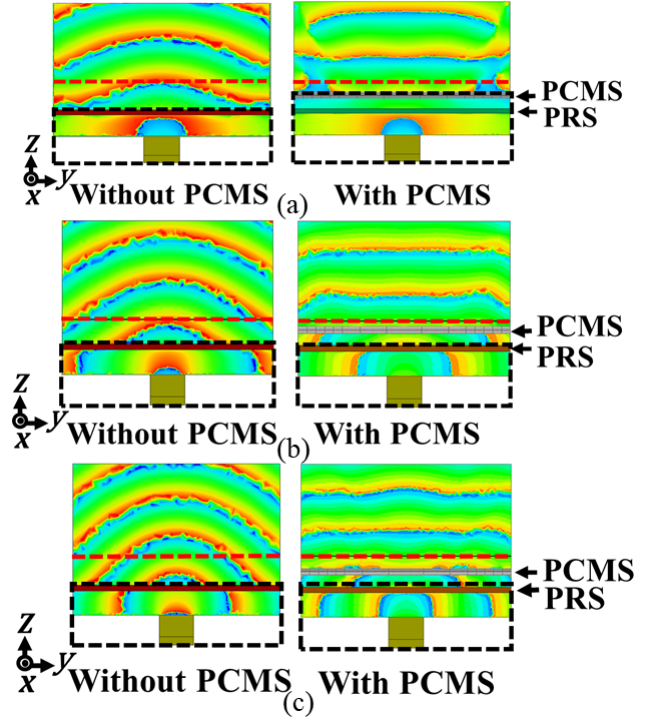


Fig. 12. Visualized comparison of propagating $\angle E_x$ in the H-plane at (a) 13.3 GHz, (b) 15 GHz, (c) 17 GHz. The red line indicates the location of 1-D probed phases in Fig. 11.

$z = 0.5\lambda_0$ from the ground plane with the help of 3D printed support and then placing the PCMS at $z = 0.25\lambda_0$ above the PRS with nylon spacers as shown in Fig. 13(b). The overall vertical height of the antenna is $0.99\lambda_0$, where λ_0 is the wavelength at the resonant frequency. We excite the antenna with a WR-62 waveguide-fed slot, with the slot dimensions of $7.9 \text{ mm} \times 11.8 \text{ mm}$. Fig. 13(c) shows a photograph of the proposed FPCA with PCMS loading in a SATIMO chamber taken during the antenna gain measurements.

We measure the $|S_{11}|$ of the proposed antenna and compare it with the simulated results in Fig. 14. It is observable that the measured $|S_{11}|$ is less than 10 dB from 12.8 GHz to

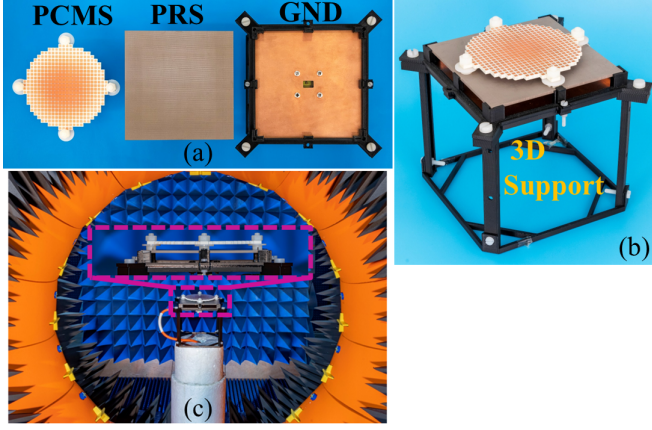


Fig. 13. (a) Photograph of the fabricated proposed antenna with PCMS; The PRS, PCMS, and ground plane are separately shown for better clarity. (b) Fully assembled fabricated prototype of proposed antenna with PCMS. (c) The proposed antenna is under testing.

19 GHz, indicating good impedance matching over a wide bandwidth of approximately 39%. Compared with the $|S_{11}|$ of the antenna without the PCMS, the integration of PCMS introduces multiple resonances in $|S_{11}|$ due to the small reflections from the PCMS unit cells which slightly impact the matching of the proposed antenna. Overall, the simulated and measured ($|S_{11}|$) agree well.

Fig. 15 compares the simulated and measured realized gains of the proposed antenna with and without the PCMS. The FPCA without the PCMS has a peak gain of approximately 12.4 dB, and the gain significantly drops away from the resonance. With the inclusion of PCMS, the measured peak realized gain is approximately 20.1 dB near 15 GHz, and the 3dB gain bandwidth ranges approximately from 13 GHz to 18.50 GHz, resulting in a relative bandwidth of 35%, representing a remarkable enhancement in comparison to previously reported PCMS FPCAs [30], [32]–[43]. Moreover, we estimate the antenna's directivity from the measured antenna gain and the simulated radiation efficiency and plot it in Fig. 15. The peak estimated directivity is approximately 20.3 dB at around 16 GHz, and the 3dB directivity bandwidth spans 36%, ranging from 13 GHz to 18.7 GHz. The gain and directivity closely follow across the operating band, with an average deviation of only 0.54 dB and less than 0.35 dB near 15.5 GHz, confirming low loss. Thus, the proposed PCMS significantly improves the conventional FPCA peak gain and the gain bandwidth. The peak aperture efficiency of our proposed design is 23%, representing an improvement over the reference antenna's efficiency of 5%. Conventional FPCAs can be optimized to maximize aperture efficiency, achieving aperture efficiencies exceeding 50%; whereas the proposed design is optimized for broadband gain performance. As shown in Figs. 3 and 5, the dominant leaky-wave field gradually attenuates as it propagates across the aperture due to continuous radiation. Therefore, we deliberately chose a larger aperture for both the reference and proposed antennas $—5.32\lambda_0 \times 5.32\lambda_0—$ to allow the leaky waves to attenuate substantially before reaching the aperture boundary. This design choice improves the broadband gain performance at the

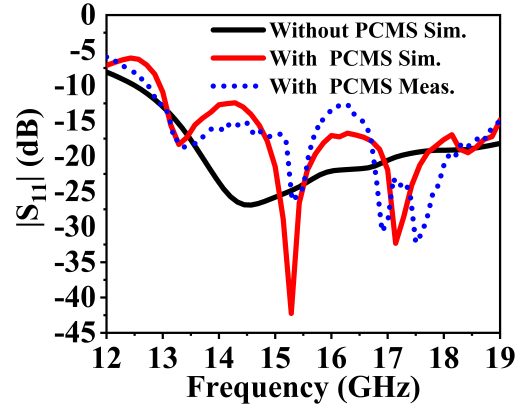


Fig. 14. The magnitude of input reflection coefficients of the proposed antenna with and without the PCMS.

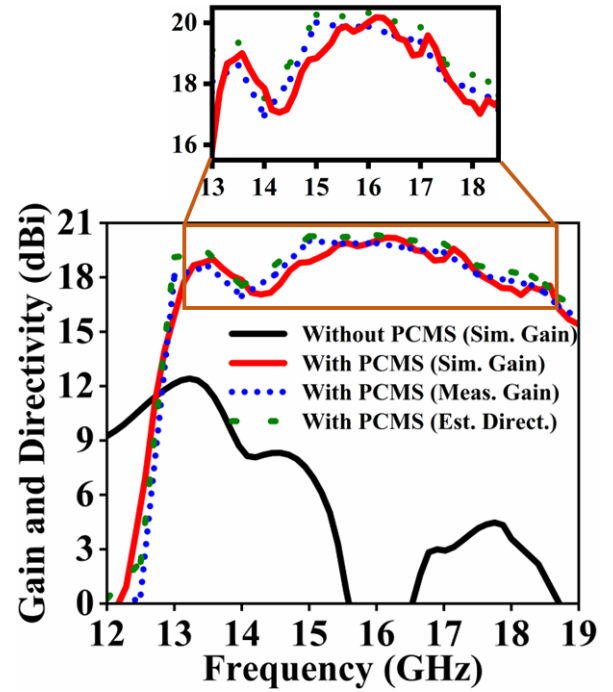


Fig. 15. Simulated and measured realized gain, along with the estimated directivity, of the proposed antenna with the PCMS. The estimated directivity is calculated separately, using the measured gain and the simulated radiation efficiency.

expense of a lower calculated aperture efficiency.

Fig. 16 compares the proposed antenna's simulated and measured radiation patterns in two principal planes. We plot the radiation patterns at five frequencies to showcase the stable radiation patterns in a wideband. It can be seen that the simulation and experimental results agree well, and the antenna has a main lobe in the broadside direction, while the sidelobes remain below -10 dB in the wideband in both E and H planes. Also, the measured cross-polarization is lower than -17 dB. Overall, the measurement results of the proposed antenna closely follow the simulated results. The minor discrepancies observed in S_{11} , gain, and radiation patterns mainly arise from practical implementation and measurement uncertainties. These include fabrication tolerances of the multilayer PCB,

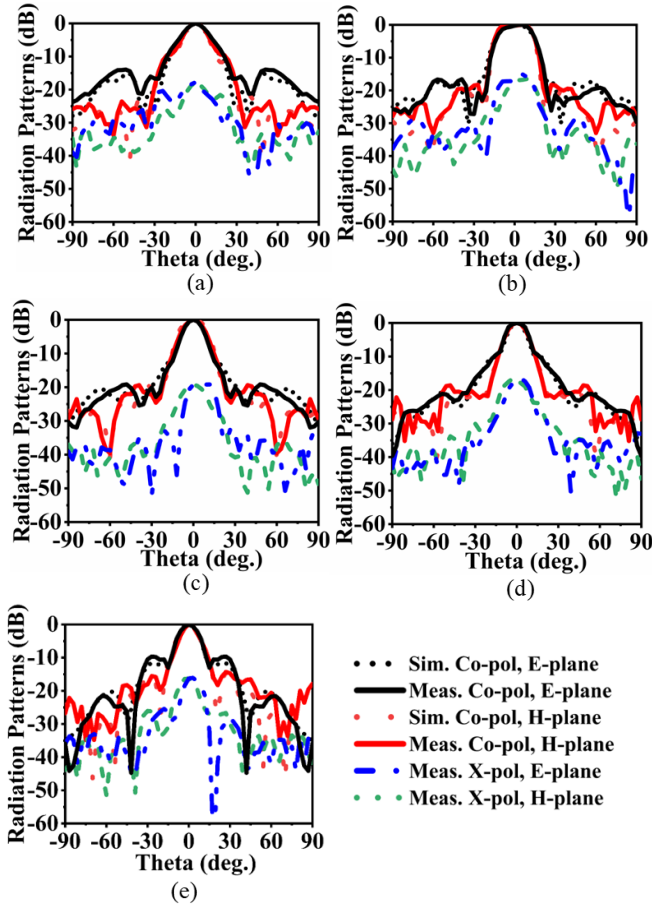


Fig. 16. Simulated and measured radiation patterns of the proposed antenna with PCMS: (a) 13.3 GHz. (b) 14.3 GHz. (c) 15.3 GHz. (d) 16.3 GHz. (e) 17.3 GHz.

dielectric loading introduced by the 3D-printed supporting structure and nylon spacers, and slight assembly and alignment errors of the PRS and PCMS. Collectively, these factors account for the small differences between the simulated and measured results.

TABLE IV
COMPARISON OF FPCA-PCMS ANTENNA DESIGNS

Ref.	Type	Area (λ_l^2)	Height (λ_0)	Gain (dBi)	3dB BW (%)	GBP	GBP/A
[32]	Dielectric	32.76	2.30	20.6	8	9.18	0.28
[33]	Printed	32.76	1.06	19.3	6	5.11	0.16
[36]	Printed	11.46	1.00	20.78	~3.4	4.07	0.36
[38]	Printed	32.76	1.01	20.8	9.4	11.30	0.35
[42]	Printed	71.15	1.16	25.6	10	36.31	0.51
[39]	Metal	31.54	1.50	21.1	12	15.46	0.49
[41]	Metal	21.80	1.65	19.4	~7.1	6.18	0.28
[58]	Metal	35.40	3.00	17.0	5.7	2.86	0.08
This Work	Printed	27.45	0.99	20.1	34.9	35.70	1.30

Table IV compares our work with other reported works. In this comparison, we consider FPCAs with PCMS designs made of dielectric, metallic, and PCB structures. This comparison clearly indicates that the proposed FPCA with printed metasurface provides the highest relative 3dB gain bandwidth while maintaining an overall compact antenna profile. The maximum peak gain of the proposed antenna is comparable to

or higher than that of the other referenced antennas. Although the gain reported in Ref. [39] is higher than that of our antenna, its electrical aperture area is significantly larger than in our case. The gain and bandwidth tradeoff of the antenna is often assessed via the gain-bandwidth product (GBP). Furthermore, comparisons that also consider the area of the antenna can be facilitated by the gain-bandwidth product per unit aperture area (GBP/A). From the table, it is evident that the antenna introduced in this paper achieves a GBP of 35.70, which is close to the best among the comparison works, and a GBP/A of 1.30, which is more than double that of the comparison works. Therefore, overall, our proposed antenna offers a compact profile, a small electrical size, comparable peak gain, a wide 3-dB gain bandwidth, and a high GBP/A, making it an effective solution for modern communication systems.

V. CONCLUSION

We demonstrate for the first time that the near-field phases of conventional Fabry-Pérot cavity antennas (FPCAs) vary quasi-linearly with frequency due to the dispersion of the dominant leaky-wave cavity modes. By analyzing the quasi-TEM mode and the first pair of leaky-wave modes, we show that the dominant leaky-wave mode governs the aperture phase behavior, while the TEM mode contributes negligibly. Building on this insight, we develop a new broadband phase-correction methodology for transmission-mode FPCAs. We design an ultrathin phase-correcting metasurface (PCMS) whose unit cells provide a phase response that varies linearly with frequency, compensating for the non-uniform aperture phase distribution. Unlike previous designs, our PCMS introduces minimal additional thickness, maintaining a total cavity height of approximately $0.99\lambda_0$ at resonance. We validate the proposed antenna through simulations and measurements, obtaining excellent agreement. The antenna achieves a peak realized gain of 20.1 dBi, a 35% 3-dB gain bandwidth, and a gain-bandwidth product per unit aperture area (GBP/A) of 1.30, indicating excellent performance and efficient aperture utilization. Crucially, the bandwidth exceeds that of previously reported planar PCMS designs by more than a factor of three. These results establish a compact and practical solution for broadband gain enhancement in transmission-mode FPCAs, with strong potential for wireless communication and imaging applications requiring wide bandwidth and high gain.

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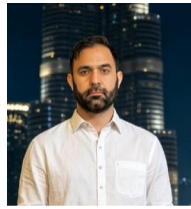
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