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A Novel Ku-Band Waveguide Piezo-Electric Air Gap Capacitor Using a Single Tuning Element

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Received: 14 May 2024 | **Revised:** 15 May 2025 | **Accepted:** 21 May 2025

Handling Editor: Yongshik Lee

Funding: The work was funded by Cranfield University internally.

Keywords: computational electromagnetics | electromagnetic wave propagation | micromechanical devices

ABSTRACT

In this work, a tuning element with controllable air gap integrated and implemented in a rectangular Ku-band WR62 waveguide is presented. The proposed tuning element concept, consists of two, top and bottom, thin conductive parallelepiped arms, which are placed in the middle of the rectangular waveguide structure. The bottom arm is bending to control the gap width (w) between the two arms, whereas the second arm is fixed. The curvature of the bending arm is controlled by a piezo-electric actuator, which affects the phase shift of the re-scattered E-field at the output port. The tuning element was designed with a 5% bandwidth centred at 15 GHz. The compact size, low-cost and the easy-to-manufacture proposed design offers a considerable phase shift with very low insertion loss, given its electrical size and operational waveband. A prototype, for the distinct frequency band, has been manufactured and measured. The same prototype has been simulated in ANSYS HFSS. The numerical results will be later used to validate the actual electromechanical prototype.

1 | Introduction

Tunable phase shifters and filters are widely utilised in microwave sensing and radio communication systems. They enable antenna beamforming [1–3] with low losses and wide phase shifting range. Mechanically reconfigurable phase shifters with tuning screws and variable loads [4–6] can effectively control the phase shifting, while moving and adjustable side walls [7, 8] can also be used to achieve the desired phase shift at the output port of a waveguide. PIN diodes can be also used [9–11] to achieve phase shift but with high losses at high frequencies. Reconfigurable materials such as liquid crystals [12, 13] and ferroelectric (paraelectric) thin films [14–16] are currently attractive for phase shifters. However, there are disadvantages, such as complex implementation within waveguide structures, and high insertion losses incurred as frequency increases.

The use of piezo-electric actuators in tuning and phase shifting techniques is emerging in the recent years as they feature of low losses and low power consumption. Coplanar waveguide (CPW) structures using piezo-electric actuators and RF MEMS have been studied recently [17, 18] to phase shift the output signal. In CPW geometries, an electrode with a piezo-electric layer [19, 20] is placed under the ground plane of the substrate ground layer and the substrate. The effective permittivity of the CPW structure is controlled electrically by this air gap and can shift the phase.

Low loss dielectric waveguide phase shifters have been implemented in microwave and millimetric wave frequencies [21, 22]. Dielectric materials with low losses show that their influence on even a small air gap between them and the waveguide wide wall can derive to a large phase shift. Reconfigurable phase shifters based on gap-waveguide technology can also be used [23]. The

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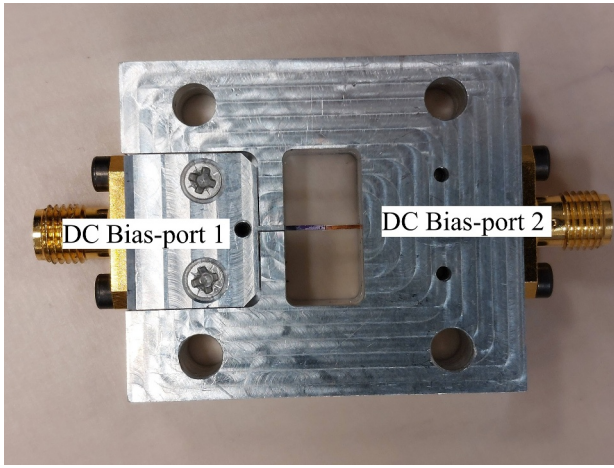


FIGURE 1 | Tuning element prototype with the two DC bias-ports.

reconfigurable behaviour is achieved by inserting a flexible metallic strip in the waveguide side wall. The variation of the metallic strip curvature, which is controlled by a tuning screw, can shorten the wide side of the waveguide and thus modify the propagation constant, producing an adjustable phase shift.

In this paper, a tuning element with controllable air gap integrated and implemented in a rectangular Ku-band WR62 waveguide is presented (Figure 1). This is a compact, low-cost, low profile, easy to manufacture design that can easily enclosed within a waveguide and provides a considerable phase shift with very low loss, given its electrical size.

2 | Tuning Element Design

A tuning element with two arms was modelled in HFSS, each arm having a certain layout as can be seen in Figure 2a. Each arm consists of nine layers of different materials and different thicknesses. Both arms are identical and are placed in a metallic cavity and stabilised by a metal flange as shown in Figure 2b. The overall bandwidth of the tuning element is 14.625--15.375 GHz (5% centred at 15 GHz).

The tuning element is placed longitudinally in the middle of the WR62 waveguide structure. The bottom arm (flexing piezo-electric actuator) of the tuning element is bending by applying a bias voltage (V) and can control the gap width (w) between the two arms. The nominal gap width (w) between the two arms is 0.2 mm.

As the applied bias voltage (V) increases, the bending of the bottom arm becomes larger and the gap width (w) between the two arms of the piezo-electric actuator increases. Thus, the phase and the propagation constant of the scattered electric field from the bending bottom arm change accordingly.

3 | Measured and Simulated Results

For the measurement setup, the DC Bias-port 1 of the DUT is negatively biased with voltage active and varying, whereas the

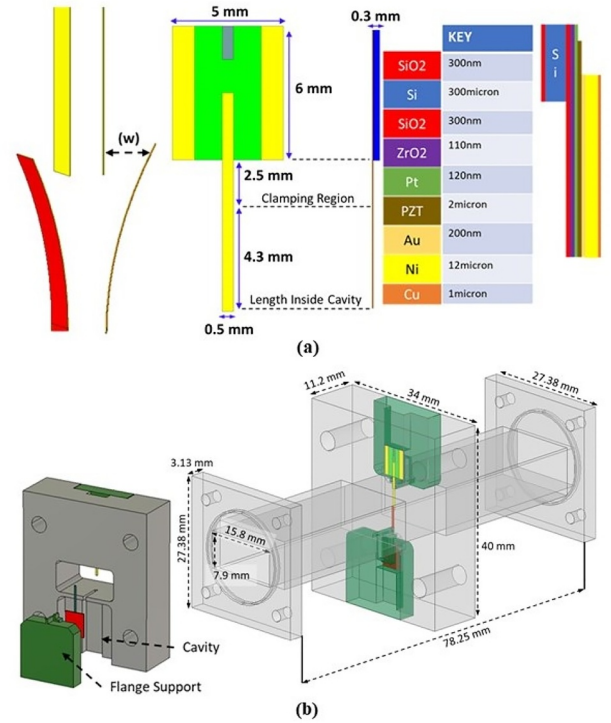


FIGURE 2 | (a) Tuning element structure layout, (b) tuning element and waveguide integration.

DC Bias-port 2 is negatively biased with the voltage active and not varying. A multiport DC source TTI EX752M is used to bias the DC DUT ports. The VNA (Anritsu 37397C) RF ports are both connected to the waveguide structure (Figure 3). Measured and simulated S-parameter results versus frequency for different DC bias values are presented. In Figure 4, the measured and simulated S_{11} and S_{21} magnitude and phase results are in a good agreement.

The S_{11} magnitude for both measured and simulated results varies from -0.15 dB to -0.7 dB when the DC bias voltage increases from 0 to 20 V. The S_{11} phase shows a good agreement between measured and simulated results with a minor difference of 20° at 14 GHz and an even smaller difference of 5° at 16 GHz. The S_{21} magnitude shows a frequency shift between each DC bias voltage state starting from 14.58 GHz, when the DC bias is 0 V and ends up to 15.5 GHz and when the DC bias is 20 V. With regards to the S_{21} phase, there is a reasonable agreement between measured and simulated results.

S-parameters are presented (Figure 5) versus the DC bias voltage, which increases from 0 to 20 V with an increasing step of 5 V. The bottom flexing piezo-electric arm of the tuning element is bending due to the applied bias voltage and increasing the gap width between the two arms. A very good agreement between measured and simulated results can be seen in Figure 5. The S_{11} Δ Phase measured and numerical results at 14, 15 and 16 GHz are 11.85, 11.91 and 8.45° and 10.7, 10.4 and 8.53° , respectively. The S_{21} Δ Phase measured results at 14, 15 and 16 GHz are 9.67, 153.33 and 7.02° , respectively. The S_{21} Δ Phase simulated results for the respective frequencies are 11.33, 39.89 and 12.7° .

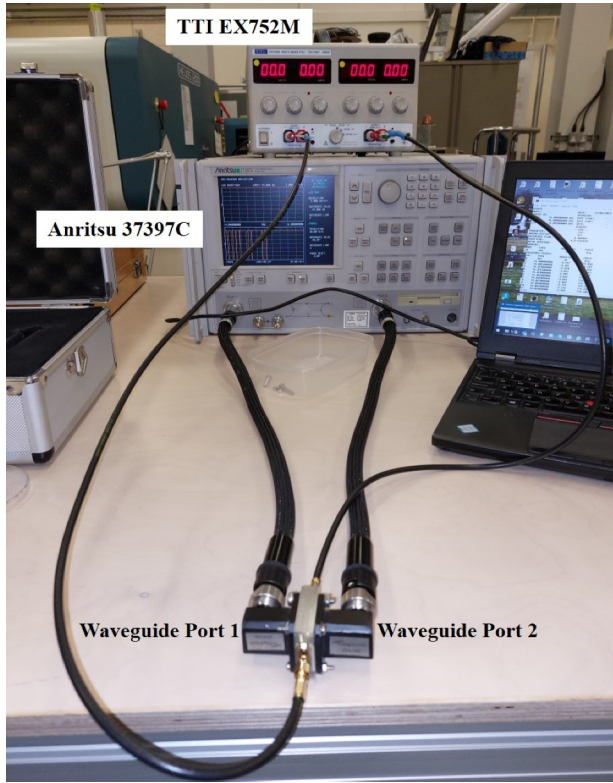
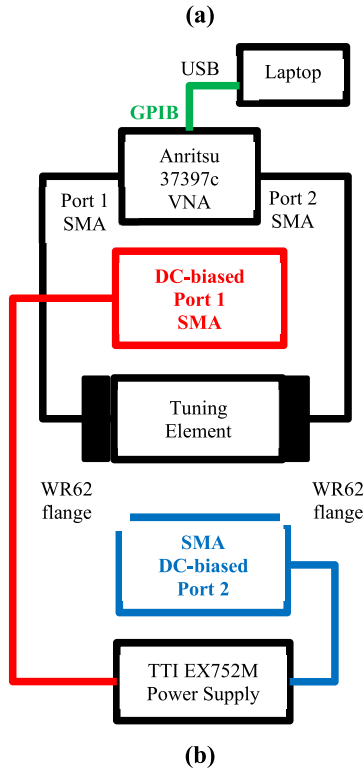


FIGURE 3 | (a) Measurement setup schematic, (b) actual measurement setup of the prototype DUT.

Table 1 summarises and compares the performance of this concept with other recently published X and Ku-band waveguide, planar and CPW topologies in terms of various aspects and factors.

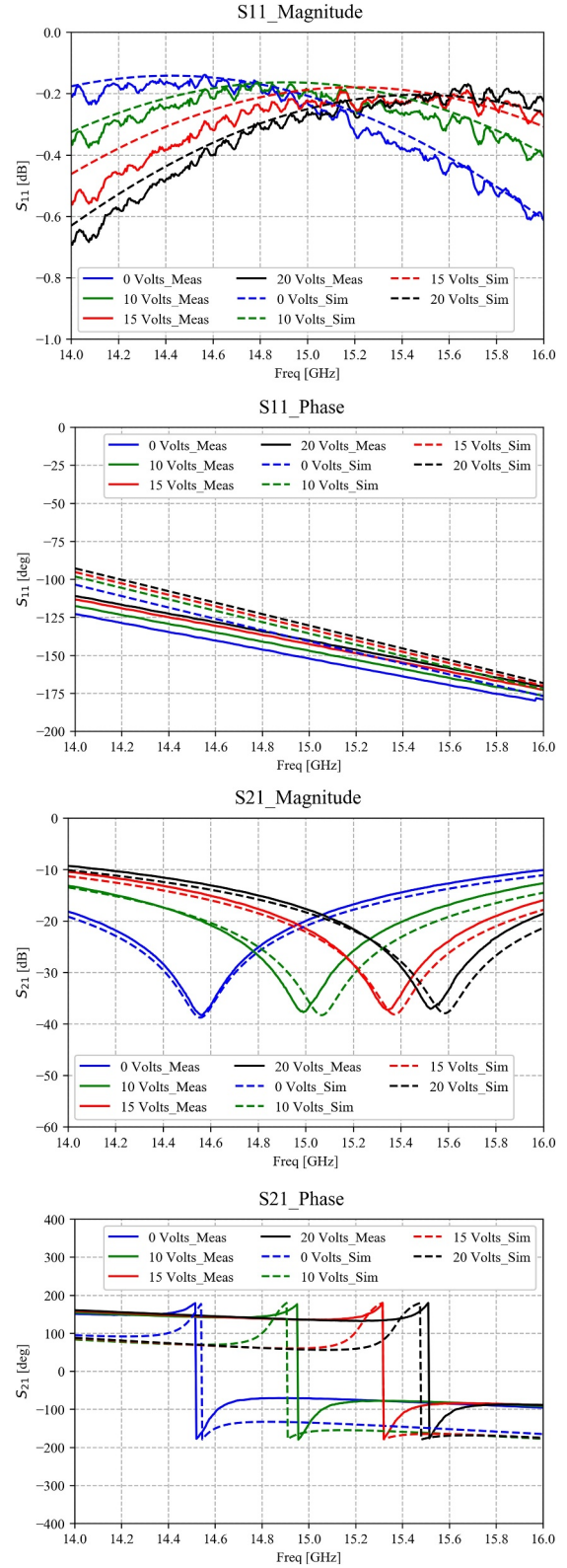


FIGURE 4 | Simulated and measured S_{11} and S_{21} magnitude (dB) and phase (deg) versus frequency (GHz) results.

3.1 | Two Port Losses Analysis and Calculations

Fundamentally, S_{11} is the amount of the wave energy leaving Port 1 when Port 2 is terminated with a 50- Ω load. S_{21} is the

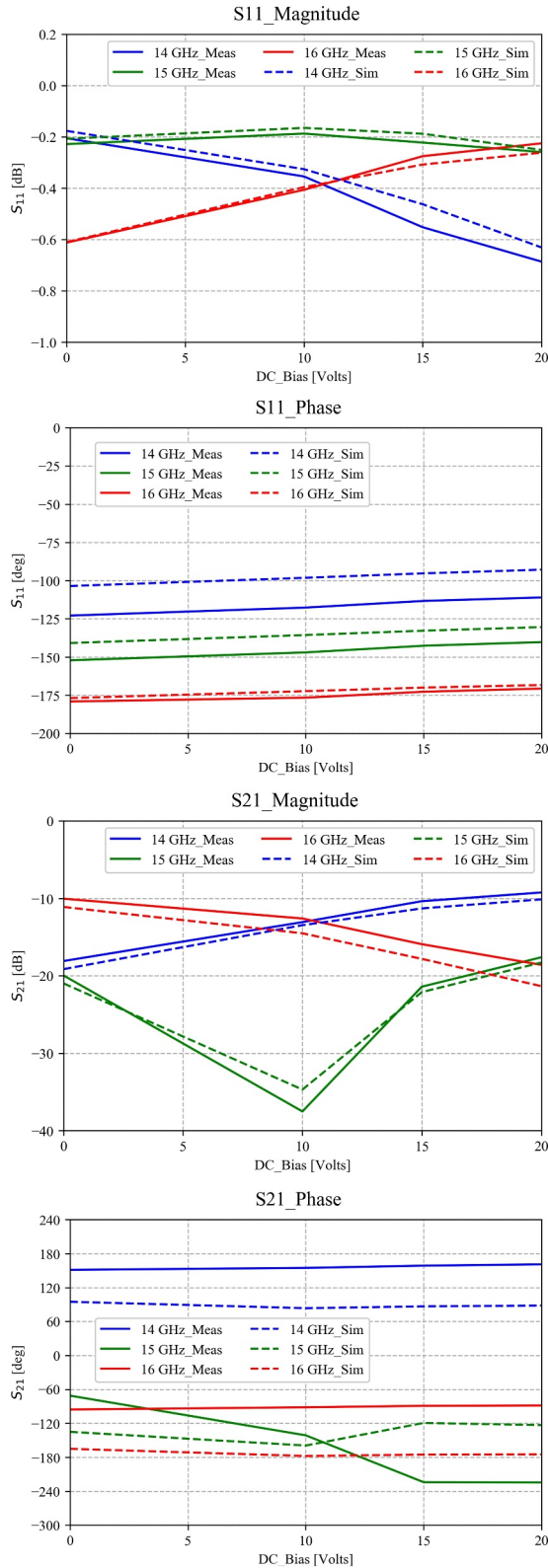


FIGURE 5 | Simulated and measured S_{11} and S_{21} magnitude (dB) and phase (deg) versus DC Bias (V) results.

energy leaving Port 2 due to an incident wave energy on Port 1. Therefore, the power loss of a two-port system is

$$1 - |S_{11}|^2 - |S_{21}|^2 \quad (1)$$

A passive lossless two-port system implies that $|S_{11}|^2 + |S_{21}|^2 = 1$. The transmission losses in dB are

$$TL[\text{dB}] = 10 \log(|S_{11}|^2 + |S_{21}|^2) \quad (2)$$

The measured and simulated transmission losses versus DC bias voltage calculated using Equation (2) can be seen in Figure 6. The minimum calculated values of the transmission losses based on measured and numerical results at 14, 15 and 16 GHz are shown in Table 2. From the calculated results, it can be seen that TL is very low for the three frequencies with reasonable agreement between the measured and simulated results.

3.2 | Predicting the Reflection Phase for a Tuning Element

The two-port S-parameter data can be used to determine the normalised lumped equivalent shunt susceptance for the tuning element, B_{NORM} . When the tuning element is placed in the waveguide with a backshort at a distance L behind it, as shown in Figure 7.

The phase of the reflected wave can be determined from normalised lumped equivalent shunt susceptance B_{NORM} and distance L . This dependence is derived from transmission line theory. This method is then applied to some acquired results for a quarter wave ($\lambda_0/4$) displaced backshort and for a further displaced case that achieves an increased phase shift as described.

In this derivation;

- a = Width of waveguide
- L = Distance from the tuning element to the backshort
- B_{NORM} = Normalised susceptance of the tuning element
- B_{BS} = Normalised susceptance backshort (short circuit at the back side of the waveguide) referred to the plane of the tuning element.

The phase of the wave reflected from a shunt susceptance is φ . The tuning element and the backshort present a combined shunt susceptance at the plane of the tuning element of $(B_{\text{NORM}} + B_{\text{BS}})$. It will be shown that the reflected phase φ is related to the combined shunt susceptance according to

$$\varphi = -2 \tan^{-1}(B_{\text{NORM}} + B_{\text{BS}}) \quad (3)$$

It is helpful to consider a single shunt susceptance B , as shown in Figure 7a, placed at the end of a transmission line of characteristic line impedance Z_0 . An incident wave of amplitude V_{in} is applied and the complex amplitude of the reflected wave is then given by $V_{\text{in}} \cdot \rho$, where ρ is the reflection coefficient.

The voltage, current and admittance across the susceptance B are equal to

$$V = V_{\text{in}}(1 + \rho) \quad (4)$$

TABLE 1 | Performance comparison.

Ref.	Freq. (GHz)	Tech.	$\Phi_{\max}$ (deg)	Losses (dB)	FoM (°/dB)
Proposed	15	W/G	200.5	0.19	1055
[4]	19	W/G	250	3	83.3
[5]	17	W/G	270	1.2	225
[9]	14.25	Planar	180	4.3	41.86
[15]	15	CPW	90	2.5	36
[16]	12	CPW	285	11	26

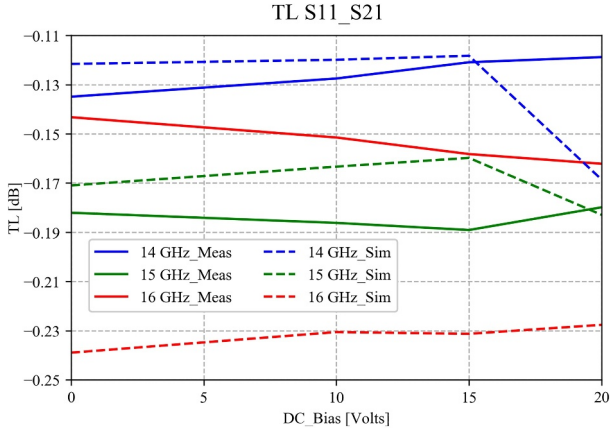

FIGURE 6 | Measured and simulated transmission losses versus DC Bias.

TABLE 2 | Measured and simulated transmission losses.

TL (dB)	14 GHz	15 GHz	16 GHz
Measured	-0.135	-0.189	-0.162
Simulated	-0.169	-0.183	-0.239

$$i = V_{in}(1 - \rho)/Z_0 \quad (5)$$

$$\frac{i}{V} = jBZ_0 = \frac{1 - \rho}{1 + \rho} \cdot \frac{1}{Z_0} \quad (6)$$

Rearranging this gives

$$\begin{aligned} (1 + \rho)jBZ_0 &= 1 - \rho \Rightarrow \rho(jBZ_0 + 1) \\ &= 1 - jBZ_0 \Rightarrow \rho = \frac{1 - jBZ_0}{1 + jBZ_0}, \Delta\rho \\ &= -2 \tan^{-1}(BZ_0) \end{aligned} \quad (7)$$

It remains to determine the normalised shunt susceptance of a backshort at a distance L towards the source of signal V_{in} as that is the plane of the tuning element to which it must be added.

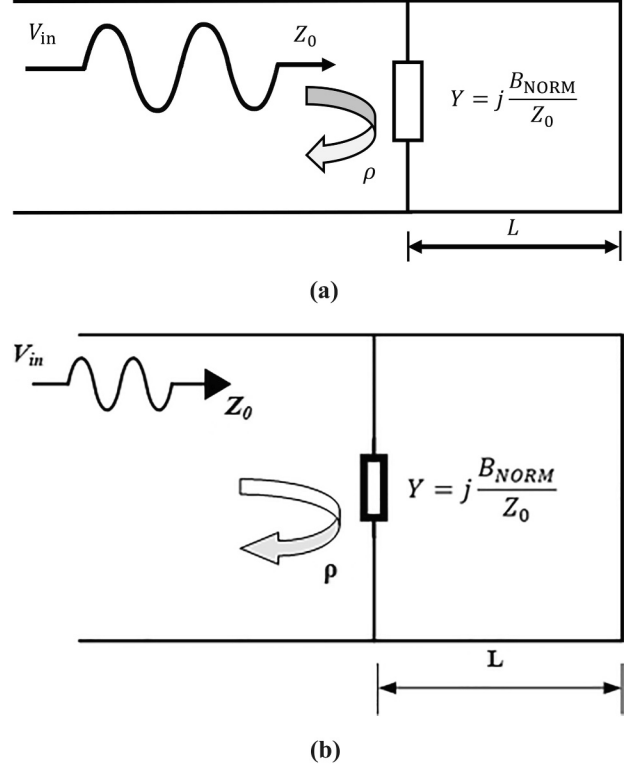
The normalised susceptance is (Z_0/j) multiplied by the absolute admittance so $B_{BS'}$ is equal to

$$\frac{1}{j}Z_0 \frac{i}{V} = \frac{1}{j} \cdot \frac{Z_0}{Z_0} \cdot \frac{V_{in}(1 - \dot{\rho})}{V_{in}(1 + \dot{\rho})} = -j \frac{(1 - \dot{\rho})}{(1 + \dot{\rho})} \quad (8)$$

For a short circuit, $\rho = -1$ at the plane of the short circuit. At a distance L towards the generator, the reflection coefficient is delayed in the phase by the round-trip phase length, therefore,:

$\dot{\rho} = \rho e^{-(2L/\lambda_g)2\pi j}$, where the wavelength λ_g in the waveguide is given by

$$\frac{1}{\lambda_g^2} = \frac{1}{\lambda_0^2} - \frac{1}{(2a)^2} \text{ where } \lambda_0 = \frac{c_0}{f} \quad (9)$$


FIGURE 7 | (a) The tuning element as a reflection phase shifter with a backshort termination displaced by L , (b) equivalent circuit of the waveguide structure with the single tuning element.

and where c_0 is the speed of light, f is the frequency and λ_0 is the wavelength in free space.

The normalised equivalent susceptance $B_{BS'}$ of the short circuit referred to the plane of the tuning element is then equal to

$$\begin{aligned} B_{BS'} &= \frac{-j(1 + e^{-2\pi j 2L/\lambda_g})}{(1 - e^{-2\pi j 2L/\lambda_g})} = -j \frac{-j2}{2j} \cdot \frac{\cos(2\pi L/\lambda_g)}{\sin(2\pi L/\lambda_g)} \Rightarrow B_{BS'} \\ &= -\cot(2\pi L/\lambda_g) \end{aligned} \quad (10)$$

The phase φ of the reflection from the combination of the tuning element and the backshort can then be calculated based on Equations (3) and (10) and it is given by the equation below:

$$\varphi = -2 \tan^{-1}[B_{NORM} - \cot(2\pi L/\lambda_g)] \quad (11)$$

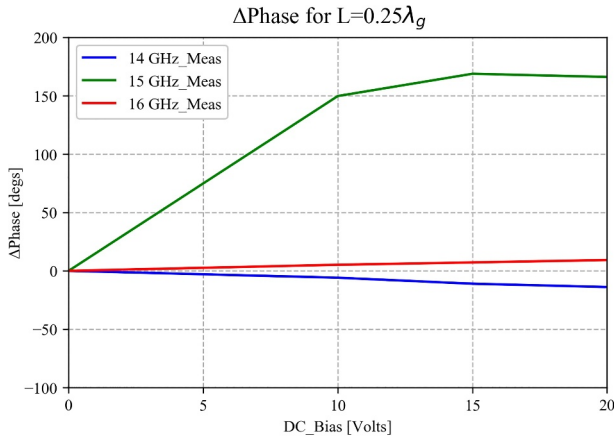


FIGURE 8 | Phase shift versus DC bias for L equal to $0.25\lambda_g$.

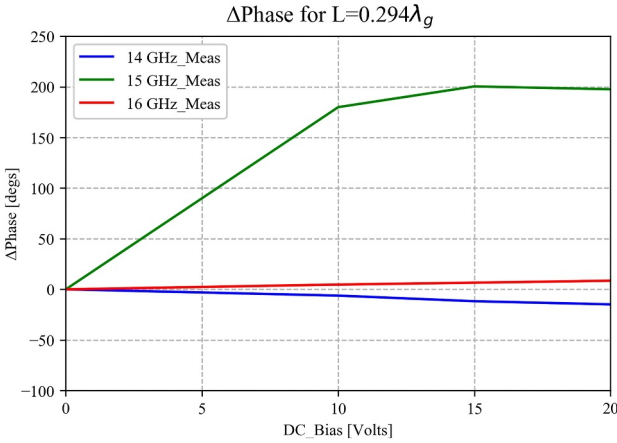


FIGURE 9 | Phase shift versus DC bias for L equal to $0.294\lambda_g$.

Figures 8 and 9 show the variation of the phase shift, whereas the backshort termination is displaced by the distance L of $0.25\lambda_g$ and $0.294\lambda_g$. The distance of the backshort termination was optimised from $0.25\lambda_g$ to $0.294\lambda_g$ and the obtained phase shift at Ku, whereas the DC voltage becomes 15 V, and is 200.5° (Table 3). The achieved phase shift at 15 GHz, when the backshort termination is placed at a distance of $0.25\lambda_g$, is 168.8° (Table 4).

On Table 3, while the DC bias varies from 0 V to 20 V and L is equal to $0.294\lambda_g$, the maximum phase shift for 14 and 16 GHz is 14.8 and 8.52° respectively. On Table 4, for the same variation of DC bias and L equal to $0.25\lambda_g$, the maximum phase shift for 14 and 16 GHz is 13.8 and 9.27° respectively.

4 | Conclusion

In this letter, a novel Ku-band waveguide piezo-electric air gap capacitor with a single tuning element was designed, fabricated, validated and measured. The low-cost configuration was easily fabricated and realised, providing a measured phase shift of more than 200° at 15 GHz with -0.19 dB transmission losses.

The proposed compact concept was modelled, simulated and optimised originally in the Ansys HFSS electromagnetic (EM)

TABLE 3 | Phase shift values when L is equal to $0.294\lambda_g$.

$L = 0.294\lambda_g$	0 V	10 V	15 V	20 V
14 GHz	0°	6.17°	11.7°	14.8°
15 GHz	0°	179.9°	200.5°	197.6°
16 GHz	0°	4.75°	6.59°	8.52°

TABLE 4 | Phase shift values when L is equal to $0.25\lambda_g$.

$L = 0.25\lambda_g$	0 V	10 V	15 V	20 V
14 GHz	0°	5.87°	11.02°	13.8°
15 GHz	0°	149.8°	168.8°	166.1°
16 GHz	0°	5.21°	7.21°	9.27°

solver giving very good agreement with the experimental results and providing a promising alternative fully developed in the waveguide technology solution for low loss RF/Microwave applications.

Future study and investigation of this concept will comprise the embedding of more than one tuning elements, the design optimisation of the tuning element array within the waveguide structure, the realisation and the verification of higher phase shift values with low transmission losses.

Author Contributions

Abraham Loutridis: data curation, formal analysis, investigation, software, validation, writing – original draft, writing – review and editing. **Ioannis Vagias:** conceptualization, funding acquisition, methodology, project administration, resources, supervision, writing – review and editing. **Glenn Leighton:** resources, supervision. **Chris Shaw:** investigation, methodology, resources, supervision, validation.

Conflicts of Interest

The authors declare no conflicts of interest.

Data Availability Statement

The data that support the findings of this study are available on request from the corresponding author. The data are not publicly available due to commercial restrictions.

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2025-01

Institution of Engineering and Technology (IET)

<https://doi.org/10.1049/mia2.70029>

Loutridis A, Vagias I, Leighton G, Shaw C. (2025) A novel Kuband waveguide piezoelectric air gap capacitor using a single tuning element. IET Microwaves, Antennas & Propagation, Volume 19, January - December 2025, Article number e70029