Coaxial Right/Left-Handed Transmission Line for Electronic Beam Steering in the Slotted Waveguide Antenna Stiffened Structure

Kelvin J. Nicholson, *Member, IEEE*, Wayne S. T. Rowe, *Member, IEEE*, Paul J. Callus, Kamran Ghorbani, *Member, IEEE*, and Tatsuo Itoh, *Fellow, IEEE*

***Abstract*—The slotted waveguide antenna stiffened structure (SWASS) utilizes conventional hat-stiffeners or blade stiffeners in aircraft sandwich structures as microwave waveguides. Slotted waveguide antenna arrays may therefore be integrated into the structure by machining slots through the outer skin. However, the primary mechanical load applied to the structure governs the orientation of these slotted waveguides and so dictates the antenna scan plane. This work extends the SWASS concept by demonstrating a means to achieve electronic phase shifting along the waveguide axis for the purpose of beam steering. This is accomplished by incorporating a varactor loaded coaxial composite right/left-handed transmission line into the SWASS for approximate matched tuning of the dispersion diagram about the design frequency.**

***Index Terms*—Coaxial, composite material structures, metamaterial, phase shifters, waveguide slot arrays.**

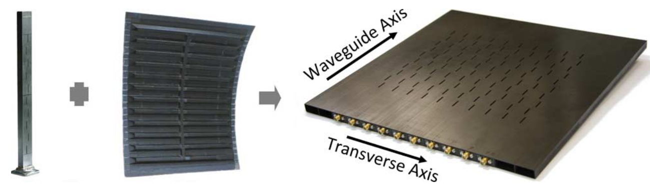

Fig. 1. SWASS concept merges the traditional slotted waveguide antenna with a typical hat-stiffened aircraft panel manufactured from aerospace CFRP.

I. Introduction

SLOTTED waveguide antennas date back to the 1940s [1] and are still in popular use today. Their mechanical robustness and simple construction favor a variety of applications across the maritime and aerospace domains. However, the increasing use of carbon fiber reinforced polymer (CFRP) has recently inspired the slotted waveguide antenna stiffened structure (SWASS) concept [2] illustrated in Fig. 1.

Previous work has addressed methods to restore the compressive strength of the SWASS (e.g., by minimizing the dimensions of the radiating slot) with negligible impact on the antenna performance [3], [4]. Unfortunately the orientation of these waveguides is governed by the primary mechanical load applied to the structure. When multiple slotted waveguides are aligned as in Fig. 1, the radiation pattern can be easily steered in the plane transverse to the waveguide axis by tuning the relative phase of each feed. However, to steer the radiation pattern in the plane parallel to the waveguide axis necessitates a method to advance or delay the propagating mode between consecutive radiating slots. Traditionally this has been achieved using in-line ferrite phase shifters, but their physical mass and size are undesirable for the SWASS concept. Hence, a new method to achieve the required in-line phase shift is proposed. This is accomplished by incorporating a coaxial composite right/left-handed transmission line (CRLH-TL) into the SWASS concept. The proposed coaxial CRLH-TL is loaded with the necessary varactor and inductor combination for approximate matched tuning of the dispersion diagram about the design frequency.

The theory and fabrication of a CRLH-TL utilizing microstrip technology is thoroughly discussed in [5]. The CRLH-TL concept has also been extended to coaxial [6] and rectangular [7] waveguides. Furthermore, a variety of methods to tune the CRLH-TL have been reported. For example, varactors were used in the microstrip implementation in [8], whereas a liquid-crystal-filled rectangular waveguide was used in [9]. Unfortunately such CRLH-TL designs are incompatible with the SWASS concept as the microstrip approach necessitates a complex bias circuit, whereas the liquid-crystal approach will significantly add to the weight of the structure. Hence, the proposed coaxial approach to the CRLH-TL is well suited to the SWASS concept since it can be tuned with a single dc bias and does not significantly add to the weight of the structure.

Recent leaky wave antenna designs have achieved fixed frequency beam steering by tuning the propagating mode along the waveguide. For example, a parallel-plate waveguide employing a varactor-loaded high impedance surface was developed in [10] for the purpose of fixed frequency beam steering. Unfortunately, only a *right*-handed mode was excited in the waveguide and so only forward radiation was permitted. To overcome this limitation, the authors later proposed a dual-feed arrangement in

Manuscript received October 13, 2013; revised January 13, 2014; accepted February 18, 2014. Date of publication March 06, 2014; date of current version April 02, 2014.

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Digital Object Identifier 10.1109/TMTT.2014.2308480

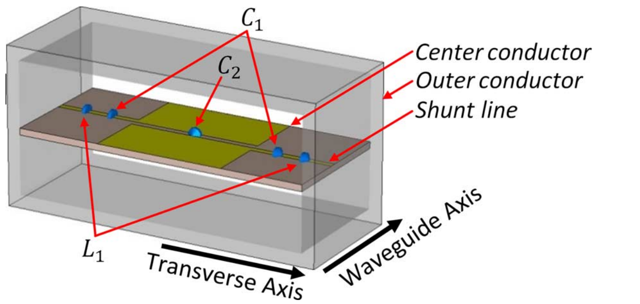

Fig. 2. Proposed unit cell of the coaxial CRLH-TL with lumped varactors (C_1 and C_2) and inductor (L_1).

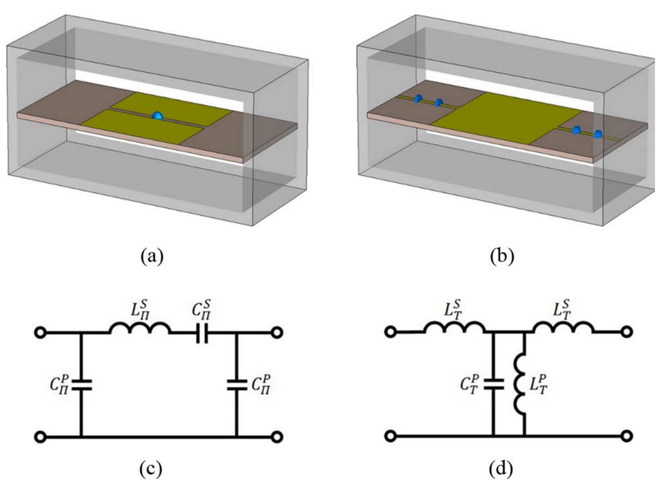

Fig. 3. Equivalent-circuit models for the proposed CRLH-TL unit cell. Simulated geometry for the: (a) $\prod$ and (b) T networks. Equivalent circuit for the: (c) $\prod$ and (d) T networks where the superscripts denote *parallel* or *series* elements.

[11] where electronic scanning from positive angles, through broad side, to negative angles was demonstrated. However, this approach to full hemispherical beam steering is inefficient. For the SWASS concept (where weight is a driving factor), the proposed CRLH-TL permits continuous tuning of the propagating mode from *right*-handed to *left*-handed at a given frequency. Fixed frequency continuous beam steering from positive angles, through broad side, to negative angles may therefore be achieved with a single feed.

II. Theory

The proposed CRLH-TL is illustrated in Fig. 2. A rectangular coaxial cross section was chosen to fit with existing SWASS panels. The center conductor and shunt lines are etched on a 0.508-mm-thick Rogers RT/Duroid 5880 substrate suspended about the mid plane of the rectangular outer conductor. Lumped circuit elements are used to model the varactors and inductors.

The dispersion diagram for the proposed transmission line can be obtained by adapting the method developed in [5]. The transmission line is regarded as the superposition of the $\prod$ and T networks illustrated in Fig. 3 with their respective equivalent circuits.

The circuit parameter C_{Π}^{S} is the parallel combination of the lumped varactor C_2 and the center conductor gap capacitance. The circuit parameter L_{T}^{P} encompasses the parallel connection of two shunt lines. Each shunt line consists of the series connection of L_1 with C_1 such that the net result remains inductive over the tunable capacitance range of the varactor. The remaining circuit parameters describe the parasitic elements present in the geometry of Fig. 2.

The circuit parameters for the Π network in Fig. 3(c) can be obtained from the simulated scattering parameters for the geometry in Fig. 3(a) according to the equations

$$C_{\Pi}^{P} = \frac{1}{j\omega}\left(Y_{11} + Y_{12}\right) \tag{1}$$

$$L_{\Pi}^{S} = \frac{-1}{2j\omega}\left[\omega\frac{\partial}{\partial\omega}\left(\frac{1}{Y_{12}}\right) + \frac{1}{Y_{12}}\right] \tag{2}$$

$$C_{\Pi}^{S} = \frac{2}{j\omega}\left[\omega\frac{\partial}{\partial\omega}\left(\frac{1}{Y_{12}}\right) - \frac{1}{Y_{12}}\right]^{-1} \tag{3}$$

where Y_{ij} are the admittance parameters and ω is the angular frequency. Similarly, the circuit parameters for the T network in Fig. 3(d) can be obtained from the simulated scattering parameters for the geometry in Fig. 3(b) according to the equations

$$L_{T}^{S} = \frac{1}{j\omega}\left(Z_{11} - Z_{12}\right) \tag{4}$$

$$C_{T}^{P} = \frac{1}{2j\omega}\left[\omega\frac{\partial}{\partial\omega}\left(\frac{1}{Z_{12}}\right) + \frac{1}{Z_{12}}\right] \tag{5}$$

$$L_{T}^{P} = \frac{2}{j\omega}\left[\frac{1}{Z_{12}} - \omega\frac{\partial}{\partial\omega}\left(\frac{1}{Z_{12}}\right)\right]^{-1} \tag{6}$$

where Z_{ij} are the impedance parameters. For this work, CST Microwave Studio was used for the numerical simulation of the geometries in Fig. 3.

Once all of the circuit parameters are known, the effective dispersion diagram of the geometry in Fig. 2 is given by the equation

$$\beta(\omega) = \sqrt{\omega^2 L_R C_R + \frac{1}{\omega^2 L_L C_L} - \left(\frac{L_R}{L_L} + \frac{C_R}{C_L}\right)} \tag{7}$$

where the per-unit-cell parameters are

$$L_R = L_{\Pi}^{S} + L_{T}^{S} \tag{8}$$

$$C_R = 2C_{\Pi}^{P} + C_{T}^{P} \tag{9}$$

$$L_L = L_{T}^{P} \tag{10}$$

$$C_L = C_{\Pi}^{S} \tag{11}$$

and the subscripts denote a contribution to either the *right*- or *left*-handed modes in the transmission line. Care must be taken to ensure the center conductor inductance (described by L_{Π}^{S} and L_{T}^{S}) is not overestimated by appropriately de-embedding the simulation ports.

The advantage of the proposed geometry is the simplicity in which the dispersion diagram may be tuned in response to a single dc-bias voltage. The outer conductor may be split such that one shunt line connects to the dc supply (V_{Bias}) while the other shunt line connects to the bias ground, as illustrated in Fig. 4(a).

A coaxial waveguide split in this manner will not significantly radiate provided the split dimension is small relative to the wavelength (a separation of 0.2 mm between V_{Bias} and

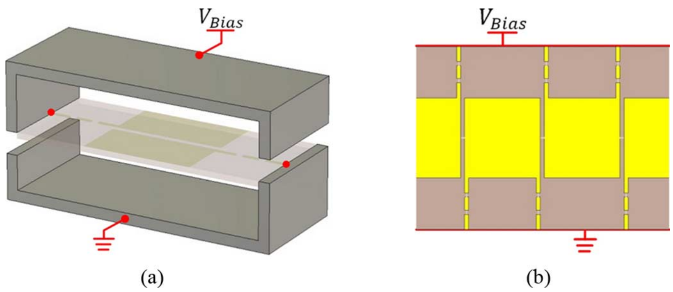

Fig. 4. Bias geometry for the proposed coaxial CRLH-TL unit cell with: (a) exaggerated split for illustration and (b) circuit board layout.

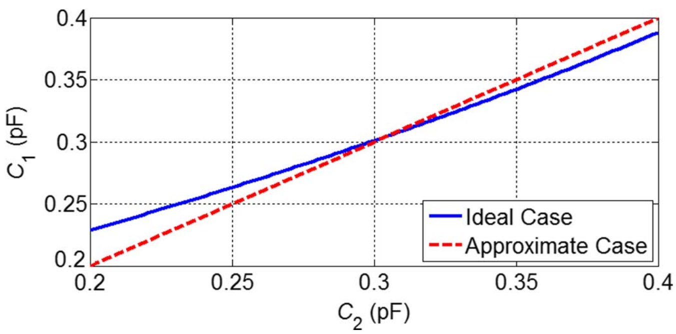

Fig. 5. Relationship between the shunt capacitance (C_1) and series capacitance (C_2) for matched tuning (blue solid line in online version) and approximate matched tuning (red dashed line in online version) at 4.0 GHz for $L_1 = 2.4$ nH and $Z = 50.0\ \Omega$.

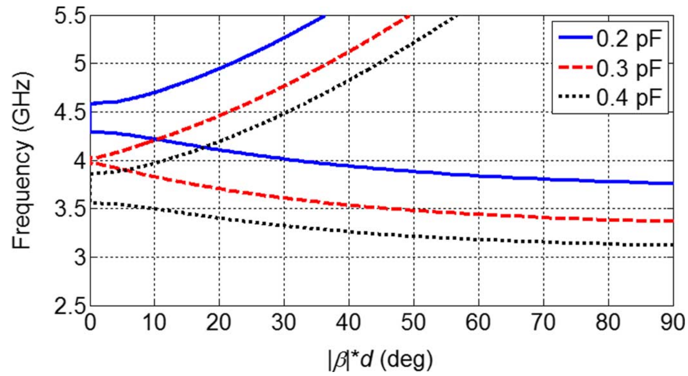

Fig. 6. Simulated dispersion when $C_1 = C_2$ is tuned over the range from 0.2 to 0.4 pF for the proposed coaxial transmission line where $d = 10.0$ mm is the unit cell length.

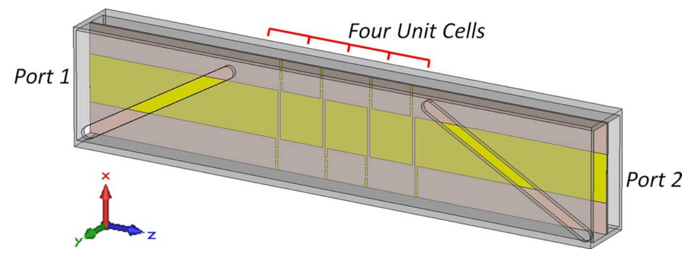

Fig. 7. Simulated geometry for a simple two-slot antenna. Four unit cells of the proposed CRLH-TL are included between the radiating slots.

ground was sufficient at 4.0 GHz for this work). Furthermore, it is easy to replicate this bias arrangement by metallizing the respective halves of the CFRP waveguides in Fig. 1. Consequently, each lumped varactor (C_1 and C_2) is subjected to the same bias voltage when every second unit cell is mirrored as in Fig. 4(b).

To simultaneously tune both the *left*- and *right*-handed modes such that no stopband separates them, it is necessary to adjust both shunt (C_1) and series (C_2) capacitances. The relationship between these capacitances for ideal matched tuning is given by the equation

$$C_1 = \frac{1}{\omega^2 (L_1 - 2L_L)} \tag{12}$$

where the angular frequency at the matched point is

$$\omega = \frac{1}{\sqrt[4]{L_R C_R L_L C_L}} \text{ with } L_L = Z^2 C_\Pi^S \tag{13}$$

and C_Π^S is a function of C_2 given a specified per-unit-cell transmission line impedance of $Z\ \Omega$. Unfortunately this precise relationship is difficult to emulate with off-the-shelf varactors. However, it can be approximated by the linear relationship $C_1 = C_2$ for the proposed transmission line, as illustrated in Fig. 5.

For this work, the outer conductor has the cross section dimensions of a standard WR-90 waveguide (22.86 mm along the transverse axis by 10.16-mm high) with a unit cell measuring $d = 10.0$ mm along the waveguide axis. The center conductor is 10.0-mm wide with 0.5-mm-wide shunt lines. The simulated dispersion diagram for this geometry is presented in Fig. 6 when $C_1 = C_2$ is tuned from 0.2 to 0.4 pF with $L_1 = 2.4$ nH.

Careful selection of the inductance $L_1 = 2.4$ nH ensures that the dispersion diagram is matched at the design frequency of 4.0 GHz when $C_1 = C_2 = 0.3$ pF. The error between the ideal case and approximate case in Fig. 5 is responsible for the stopband that appears between the *left*- and *right*-handed modes in Fig. 6 as the varactors are tuned. However, at 4.0 GHz, a propagating mode is always present. Hence, this mode can be used to achieve the necessary phase advance or delay between consecutive radiating slots. This simple addition to the SWASS concept will therefore provide the ability to steer the radiating beam in the plane parallel to the waveguide axis. A different value for L_1 would not have yielded the same matched condition at the design frequency and thus would have prevented continuous beam steering from positive angles, through broad side, to negative angles.

III. Simulation of Two Slot Steerable Antenna

To demonstrate how the proposed CRLH-TL may be applied to the SWASS concept, consider the leaky-wave antenna geometry in Fig. 7. When a quasi-TEM wave is propagating in the $+z$ direction and the structure is terminated with a matched load, the total radiated field is polarized along the x axis.

Unlike conventional leaky-wave antennas (where a continuous slot or partially reflective surface typically form the antenna aperture), the SWASS concept necessitates discrete slots to ensure some degree of structural advantage. Hence, the antenna in Fig. 7 utilized two discrete slots. The slots were cantered at 30° relative to the waveguide axis and were separated by one guided wavelength. This permitted a maximum

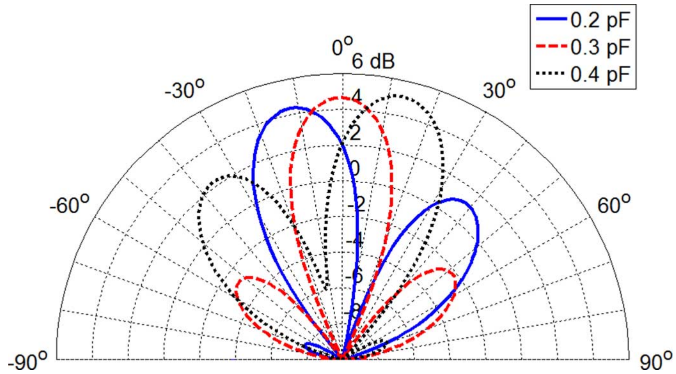

Fig. 8. Simulated realized antenna gain at 4.0 GHz in the y–z-plane for the geometry in Fig. 7.

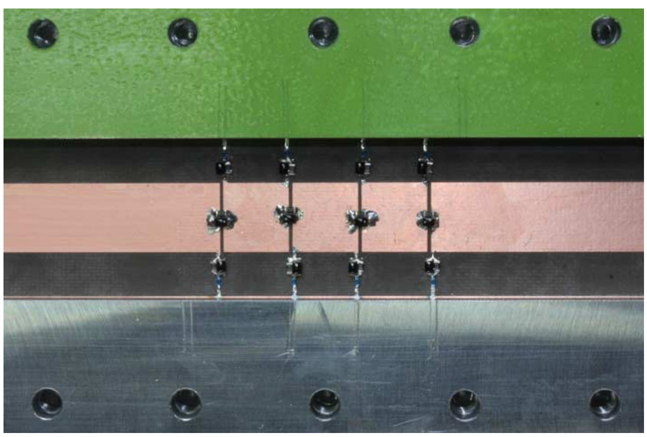

Fig. 9. Four unit cells of the proposed coaxial CRLH-TL structure with cover plate removed. The bias voltage is applied between the upper side wall (insulated with 0.2-mm-thick green tape) and the lower aluminum sidewall.

of four CRLH-TL units to be included between consecutive slots. More than four CRLH-TL units will yield a net phase advance exceeding 180° at 4.0 GHz between consecutive slots and would therefore result in a decreased realized gain. The slot width was fixed at 2.6 mm, whereas the slot lengths were 39.1 and 42.6 mm, respectively, to ensure equal radiated power at 4.0 GHz in the absence of loss in the CRLH-TL. The simulated realized gain of the proposed antenna at 4.0 GHz is illustrated in Fig. 8.

The proposed antenna exhibits a maximum scan angle of $\pm 12^{\circ}$ from broad side with peak realized gain spanning from 4.2 to 5.2 dB. Although improvements to the sidelobe level and scan angle are desirable, this antenna clearly demonstrates how the proposed coaxial CRLH-TL can be applied to the SWASS concept to achieve electronic beam steering.

IV. Measurement of CRLH-TL and Slot Antenna

Four unit cells of the proposed coaxial CRLH-TL were fabricated as illustrated in Fig. 9. A metallic test fixture was used to validate the design in the absence of the complex CFRP conductivity [12]. The MGV-125–08-0805 varactor manufactured by Aeroflex Metelics was used to implement C_1 and C_2. The inductor LQW15AN2N4B00D manufactured by Murata was used to implement L_1. To ensure equal division of the reverse bias,

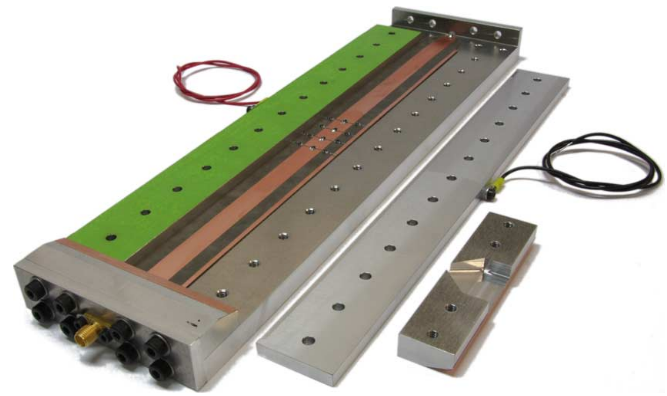

Fig. 10. Test fixture (disassembled) with tapered transition from SMA to rectangular coaxial.

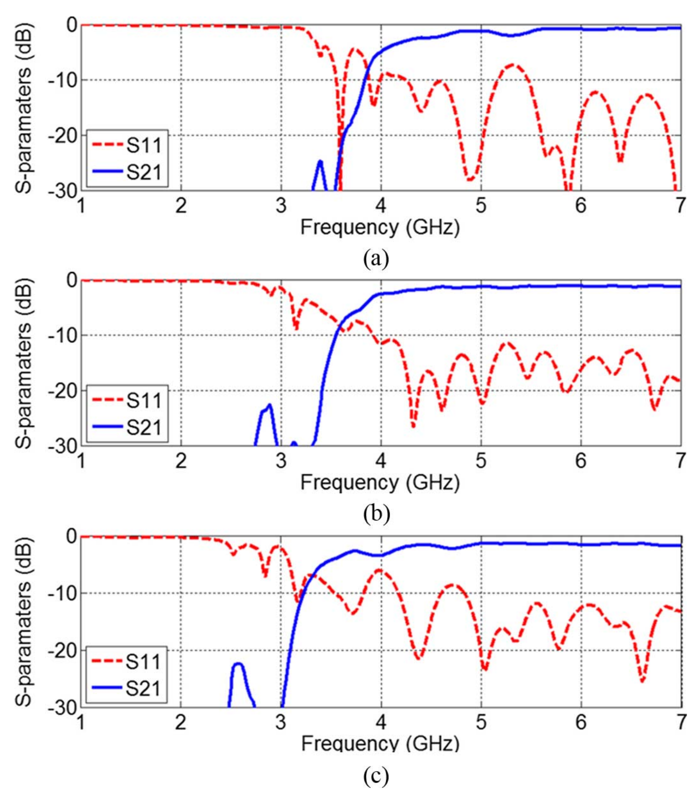

Fig. 11. Measured CRLH-TL S-parameters for a total reverse bias of: (a) 60.0 V, (b) 21.0 V, and (c) 12.0 V, which equates to: (a) 0.2 pF, (b) 0.3 pF, and (c) 0.4 pF per varactor.

a 1.0-MΩ resistor (CPF0603B1M0E by Tyco Electronics) was placed in parallel with each varactor.

To excite the quasi-TEM mode in the manufactured CRLH-TL structure, a simple tapered transition from SMA to the rectangular coaxial waveguide was designed and manufactured. The transition maintains a 50.0-Ω impedance by linearly tapering the center and outer conductor dimensions over a distance of 10.0 mm. The same transition was used at either end of the CRLH-TL. The disassembled test fixture (with no cover plate) is illustration in Fig. 10.

The S-parameters for the manufactured CRLH-TL were measured (with closed cover plate) for three reverse bias voltages, as illustrated in Fig. 11.

Fig. 12. Manufactured test fixture with slotted cover plate. Nylon screws were used to ensure the bias was not shorted once the cover plate was installed.

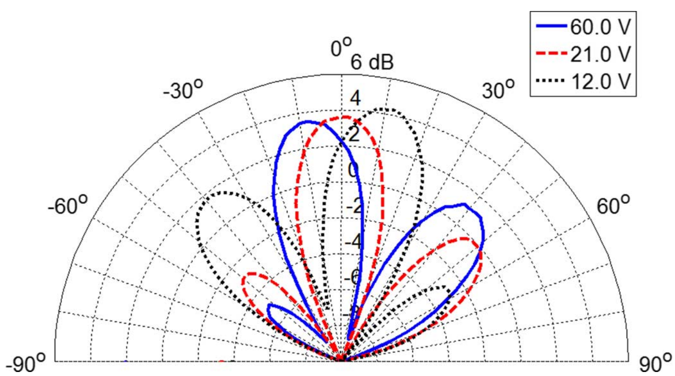

Fig. 13. Measured realized antenna gain at 4.0 GHz in the y–z-plane.

The results in Fig. 11 agree with the simulated dispersion in Fig. 6. For example, a reverse bias of 21.0 V equates to a capacitance of 0.3 pF per varactor such that both *left*- and *right*-handed modes are matched with a passband extending from approximately 3.4 GHz. For a reverse bias of 60.0 V (0.2 pF per varactor) and 12.0 V (0.4 pF per varactor), the onset of the *left*-handed mode is evident at approximately 3.7 and 3.1 GHz, respectively, as predicted in Fig. 6. However, the stopbands are not clear in the measured result since only four unit cells were included in the manufactured transmission line. Increasing the number of unit cells in the manufactured transmission line may further enhance the stopband characteristic predicted in Fig. 6.

The closed cover plate was then replaced with the slotted cover plate (as illustrated in Fig. 12) and Port 2 was terminated with a matched load.

The measured realized gain of the two-slot antenna in the y–z plane is illustrated in Fig. 13. The measured antenna exhibits a maximum scan angle of $\pm 11^{\circ}$ from broad side with peak realized gain spanning from 3.6 to 4.3 dB. Hence, the measured antenna demonstrates reasonable agreement with the simulated results in Fig. 8. However, there is slight discrepancy in gain (less than 1.0 dB) and scan angle (less than 1.0°) that is attributed to resistive losses that were not included in the simulation.

For completeness, the simulated and measured scan angle and return loss ($|S_{11}|$ when port 2 is terminated with a matched load) is presented in Fig. 14 as a function of applied bias voltage.

The results in Fig. 14(a) clearly demonstrate continuous beam steering from positive angles, through broad side, to negative angles at 4.0 GHz. Hence, the transition from *left*- to *right*-handed modes in the proposed CRLH-TL confirms the theoretical dispersion diagram in Fig. 6. Furthermore, Fig. 14(b) demonstrates that the proposed antenna is reasonably matched over the entire tunable voltage range.

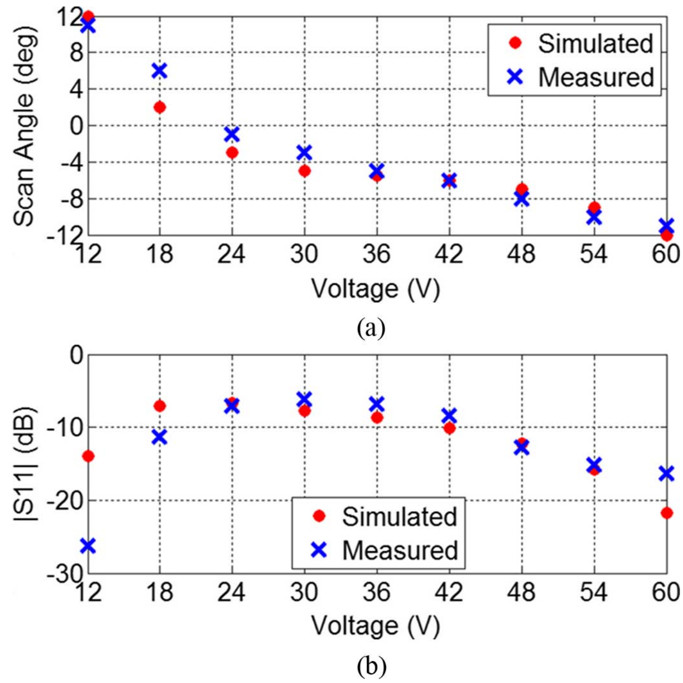

Fig. 14. Simulated and measured: (a) scan angle and (b) return loss at 4.0 GHz versus applied bias voltage.

V. Conclusion

This paper has presented a simple CRLH-TL that permits continuous electronic beam steering in the SWASS concept. Cascading the proposed unit cell such that every second cell along the transmission line is mirrored permits the approximate matched tuning of the dispersion diagram in response to a single dc-bias voltage. Consequently, a positive or negative phase shift can be achieved between consecutive radiating slots at the design frequency. This work has therefore demonstrated how electronic beam steering can be achieved with the SWASS concept regardless of the principal mechanical load orientation.

Acknowledgment

This work was conducted within the Defence Materials Technology Centre, which was established and is supported by the Australian Government's Defence Future Capability Technology Centre (DFCTC).

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