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

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.

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
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 $\Pi$ and $T$ networks illustrated in Fig. 3 with their respective equivalent circuits.

The circuit parameter $C_{\Pi}$ is the parallel combination of the lumped varactor $C_V$ and the center conductor gap capacitance. The circuit parameter $L_P$ 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 $\Pi$ network in Fig. 3(c) can be obtained from the simulated scattering parameters for the geometry in Fig. 3(a) according to the equations

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

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

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

where $Y_{ij}$ are the admittance parameters and $\omega$ 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

$$L_T = \frac{1}{j\omega} \left( Z_{11} - Z_{12} \right)$$

$$C_T = \frac{1}{2j\omega} \left[ \frac{\partial}{\partial \omega} \left( \frac{1}{Z_{12}} \right) \right] + \frac{1}{Z_{12}}$$

$$L_T = \frac{2}{j\omega} \left[ \frac{\partial}{\partial \omega} \left( \frac{1}{Z_{12}} \right) \right] - \frac{1}{Z_{12}}$$

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

$$\beta(\omega) = \sqrt{\omega^2 L_r C_R + \frac{1}{\omega^2 L_1 C_L}} - \left( \frac{L_T}{L_L} + \frac{C_R}{C_L} \right)$$

where the per-unit-cell parameters are

$$L_R = L_{\Pi} + L_T$$

$$C_R = 2C_{\Pi} + C_T$$

$$L_L = L_P$$

$$C_L = C_{\Pi}$$

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
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

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

and $C_2^{\text{SWASS}}$ is a function of $C_2$ given a specified per-unit-cell transmission line impedance of $Z_0$. 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 $z$ 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 dispersion.
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 ±12° 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, 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 illustrated 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.
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 ±11° 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 (|S11| 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.

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.

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REFERENCES

Wayne S. T. Rowe (S’98–A’01–M’04) received the B.Eng (Hons.) and Ph.D. degrees from RMIT University, Melbourne, Australia, in 1998 and 2002, respectively.

He is currently an Associate Professor with RMIT University. He has performed research and consulting work for companies including Cochlear, DSTO, CME Pty Ltd., SP Ausnet, and BAE Systems. He is a Reviewer for *IEEE Microwave, Antennas and Propagation* and *PIER*. His research interests include conformal, embedded, and load-bearing printed antennas/sensor structures, in-integrated antenna/microwave modules, tunable metamaterials, wireless power transmission, and RF sensors for high-voltage (HV) power lines.

Dr. Rowe is a reviewer for the *IEEE TRANSACTIONS ON ANTENNAS AND PROPAGATION* and the *IEEE ANTENNAS AND WIRELESS PROPAGATION LETTERS*.

Kamran Ghorbani (M’04) received the B.Eng (Hons.) and Ph.D. degree from RMIT University, Melbourne, Vic., Australia, in 1995 and 2001, respectively.

He is the Director of the Communication and Technologies Research Centre, RMIT University. He is responsible for strategic planning and managing the research center. He is a Reviewer for *Applied Physics Letters* and *Electronics Letters*. His research interests include antennas, microwave/millimeter systems, RF devices, ferroelectric materials, composite materials and phased-array antennas.

Prof. Ghorbani is a reviewer for the *IEEE Antennas and Propagation Society* and the *IEEE Microwave Theory and Techniques Society (IEEE MTT-S)*.

Tatsuo Itoh (S’69–M’69–SM’74–F’82) received the Ph.D. degree in electrical engineering from the University of Illinois at Urbana-Champaign, Urbana, IL, USA, in 1969.

From September 1966 to April 1976, he was with the Electrical Engineering Department, University of Illinois at Urbana-Champaign. From April 1976 to August 1977, he was a Senior Research Engineer with the Radio Physics Laboratory, SRI International, Menlo Park, CA, USA. From August 1977 to June 1978, he was an Associate Professor with the University of Kentucky, Lexington, KY, USA. In July 1978, he joined the faculty of The University of Texas at Austin, Austin, TX, USA, where he became a Professor of electrical engineering in 1981 and the Director of the Electrical Engineering Research Laboratory in 1984. During the summer of 1979, he was a Guest Researcher with AEG-Telefunken, Ulm, Germany. In September 1983, he was selected to hold the Hayden Head Centennial Professorship of Engineering at The University of Texas at Austin. In September 1984, he became the Associate Chairman for Research and Planning of the Electrical and Computer Engineering Department, The University of Texas at Austin. In January 1991, he joined the University of California at Los Angeles (UCLA), Los Angeles, CA, USA, as Professor of electrical engineering and Holder of the TRW Endowed Chair in Microwave and Millimeter Wave Electronics. He was an Honorary Visiting Professor with the Nanjing Institute of Technology, Nanjing, China, and with the Japan Defense Academy. In April 1994, he became an Adjunct Research Officer with the Communications Research Laboratory, Ministry of Post and Telecommunication, Tokyo, Japan. He currently holds a visiting professorship with The University of Leeds, Leeds, U.K. He has authored or coauthored 310 journal publications and 640 refereed conference presentations. He has authored 30 books/book chapters in the area of microwaves, millimeter waves, antennas, and numerical electromagnetics. He has generated 60 Ph.D. students.

Dr. Itoh is a member of the Institute of Electronics and Communication Engineers of Japan, and Commissions B and D, USNC/URSI. He was the editor-in-chief for the *IEEE TRANSACTIONS ON MICROWAVE THEORY AND TECHNIQUES* (1983–1985). He serves on the Administrative Committee of the IEEE Microwave Theory and Techniques Society (IEEE MTT-S). He was vice president of the IEEE MTT-S in 1989 and president in 1990. He was the editor-in-chief of IEEE MICROWAVE AND GUIDED WAVE LETTERS (1991–1994). He was elected an Honorary Life Member of the IEEE MTT-S in 1994. He was elected a member of the National Academy of Engineering in 2003. He was the chairman of the USNC/URSI Commission D (1988–1990) and chairman of Commission D of the International URSI (1993–1996). He is chair of the Long Range Planning Committee, URSI. He serves on advisory boards and committees for numerous organizations. He was the recipient of numerous awards including the 1998 Shida Award presented by the Japanese Ministry of Post and Telecommunications, the 1998 Japan Microwave Prize, the 2000 IEEE Third Millennium Medal, and the 2000 IEEE MTT-S Distinguished Educator Award.