The average root-mean-square error from 1 to 40 GHz for 17 different CPW lines on GaAs with a metal thickness of 1.58 μm is shown in Table I. It is seen that of the previous equations in the literature, those in [7] and [8] are the most accurate with an average error of approximately 11%, whereas an average error of 40% and a maximum error of 56% is obtained using Wheeler's incremental inductance rule. The average error for (1)–(3), shown in Table I, is only 4.7%, and if all three metal thickness are included, the average error is 6%. This is better than any of the previously reported equations, and most of this error is at low frequency where af^b does not accurately fit the data, as discussed earlier; the maximum error is 35% at 1 GHz for the thickest metal lines.

IV. Conclusions

This paper has presented a new closed-form equation for the prediction of CPW attenuation that is valid for a wide range of center strip conductor widths, slot widths, and metal thickness. It has been shown that this simple equation is as accurate as more complicated closed-form equations with an average error from 1 to 40 GHz of approximately 6%. Furthermore, the equation is invertible if one assumes f^b varies slowly with S and W to permit a determination of SW required to achieve a desired attenuation. A comparison between five previous closed-form equations for CPW attenuation to measured data has been made.

References

[1] K. C. Gupta, R. Garg, and I. J. Bahl, *Microstrip Lines and Slotlines*. Norwood, MA: Artech House, 1979, pp. 275–287.
[2] C.-L. Liao, Y.-M. Tu, J.-Y. Ke, and C. H. Chen, "Transient propagation in lossy coplanar waveguides," *IEEE Trans. Microwave Theory Tech.*, vol. 44, pp. 2605–2611, Dec. 1996.
[3] R. E. Collin, *Foundations for Microwave Engineering*. New York: McGraw-Hill, 1992, pp. 905–910.
[4] G. Ghione, "A CAD-oriented analytical model for the losses of general asymmetric coplanar lines in hybrid and monolithic MIC's," *IEEE Trans. Microwave Theory Tech.*, vol. 41, pp. 1499–1510, Sept. 1993.
[5] R. Goyal, Ed., *Monolithic Microwave Integrated Circuits: Technology & Design*. Norwood, MA: Artech House, 1989, p. 375.
[6] G. H. Owyang and T. T. Wu, "The approximate parameters of slot lines and their complement," *IRE Trans. Antennas Propagat.*, vol. AP-6, pp. 49–55, Jan. 1958.
[7] C. L. Holloway and E. F. Kuester, "A quasi-closed form expression for the conductor loss of CPW lines, with an investigation of edge shape effects," *IEEE Trans. Microwave Theory Tech.*, vol. 43, pp. 2695–2701, Dec. 1995.
[8] W. Heinrich, "Quasi-TEM description of MMIC coplanar lines including conductor-loss effects," *IEEE Trans. Microwave Theory Tech.*, vol. 41, pp. 45–52, Jan. 1993.
[9] G. E. Ponchak, E. M. Tentzeris, and L. P. B. Katehi, "Characterization of finite ground coplanar waveguide with narrow ground planes," *Int. J. Microcircuits and Elect. Pack.*, vol. 20, no. 2, pp. 167–173, 1997.
[10] R. B. Marks, "A multiline method of network analyzer calibration," *IEEE Trans. Microwave Theory Tech.*, vol. 39, pp. 1205–1215, July 1991.
[11] W. H. Haydl, J. Braunstein, T. Kitazawa, M. Schlechtweg, P. Tasker, and L. F. Eastman, "Attenuation of millimeter-wave coplanar lines on gallium arsenide and indium phosphide over the range 1–60 GHz," in *IEEE Int. Microwave Symp. Dig.*, June 1992, pp. 349–352.
[12] K. C. Gupta, R. Garg, and R. Chadha, *Computer-Aided Design of Microwave Circuits*. Norwood, MA: Artech House, 1981.

A New Phase-Shifterless Scanning Technique for a Two-Element Active Antenna Array

Young-Huang Chou and Shyh-Jong Chung

***Abstract*—In this paper, a new phase-shifterless beam-scanning technique is proposed and demonstrated for a two-element active antenna array. An internal control line with an embedded amplifier is introduced in the array to provide another injection signal other than the mutual-coupling injection signal between antennas. By mixing the effects of the signals from the control line and mutual coupling, the phase difference between antennas is adjusted by gradually changing the amplifier bias on the control line. A dynamic analysis is presented to explain the scanning mechanism. The measured results showed that, when the control-line amplifier was biased from the off state to fully on state, the radiation pattern of the array was varied smoothly from the out-of-phase mode (with 180° radiation phase difference) to the in-phase mode (with 0° phase difference). During the scanning process, the antenna oscillators were stably locked, with the deviation of the locked frequency lower than 0.35%.**

***Index Terms*—Active antenna array, phase-shifterless scattering.**

I. Introduction

In the applications of microwave and millimeter-wave systems, the spatial or quasi-optical power-combining techniques are used increasingly for high-power generation and low-loss transmission with the limitation of available power of solid-state devices [1], [2]. Active integrated antennas, which combine passive antennas with active circuits, are implemented in an array to fulfill these requirements. Each active antenna acts not only as a radiator, but also an oscillator. By means of the injection-locking signals, these oscillators can be synchronized to a common frequency and the powers radiated from the antennas can be effectively combined in free space. Two types of injection-locking distribution have been used in the literature. In the first type, the injection signals come from the mutual couplings between array elements, either through free space (weak coupling) [3], [4] or through an embedded coupling network (strong coupling) [5]. Depending on the inter-element distance, the phase delay between adjacent antennas can be locked to an in-phase mode (with 0° phase delay) or an out-of-phase mode (with 180° phase delay) [3], [5]. In the second type, the injection signals originate from an external oscillator and are distributed to all the oscillators in the array [6]. The phase delay is determined by the path-length differences of the injection signals to the oscillators, so that the array radiation field can be focused to a predesigned direction.

In addition, the beam-scanning ability for these active antenna arrays has recently become attractive. By taking the advantages of the injection-locking nonlinear effect to alter the phase delay progressively, the array radiation beams can be scanned without using expensive phase shifters [4], [5], [7]–[9]. Stephan first demonstrated the beam-scanning phenomenon by adjusting the phases of the external signals injected to the peripheral active antennas of the array [7]. Since the total phase delay from the first to last antennas was bounded by the phase difference between the two external injection signals, the beam-scanning range was thus limited and inversely

Manuscript received February 23, 1998; revised October 17, 1998. This work was supported by the National Science Council of the Republic of China under Grant NSC 87-2215-E-009-064.

The authors are with the Department of Communication Engineering, National Chiao Tung University, Hsinchu, Taiwan 300, R.O.C.

Publisher Item Identifier S 0018-9480(99)01142-4.

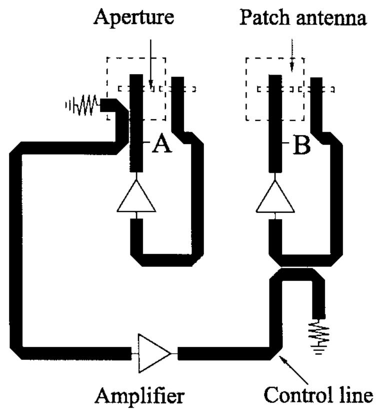

Fig. 1. A two-element scanning active arrays using feedback aperture-coupled microstrip antenna oscillators.

proportional to the number of antennas in the array. Liao and York [8] employed a different approach to improve this disadvantage. In their design, the beam scanning was accomplished by tuning the oscillating frequencies of the array peripheral elements. A nonlinear analysis for quasi-optical oscillator arrays was also introduced by York [4] to explain the beam-steering mechanism of this approach. It seemed that the steering range was limited unless the oscillators were closely located. Another method was proposed by Chew *et al.* [9], where unilateral injection locking was used between the array elements. The injection signal for the latter oscillator was tapped from the previous one and that for the first oscillator was provided from an external source. By changing the frequency of the external source, the radiation phases of the antennas were controlled one after one.

In this paper, a new phase-shifterless beam-scanning technique is proposed and demonstrated for a two-element active antenna array. An internal control line with an embedded amplifier is introduced in the array to provide another injection signal other than the mutual-coupling injection signal between antennas. By mixing the effects of the signals from the control line and the mutual coupling, the phase difference between antennas can be adjusted by gradually changing the amplifier bias on the control line. A theory modified from that in [4] is described in Section II to explain the present scanning mechanism, followed by simulation and measurement results in Section III. Section IV gives the conclusions.

II. Theory

Fig. 1 shows the proposed configuration of the scanning array, which contains two feedback aperture-coupled microstrip antenna oscillators [10] and an extra control line with an embedded amplifier of variable gain. The two-layer antenna oscillator was designed on an antenna substrate of $\varepsilon_r = 2.33$, $h = 31$ mil (thickness) and a circuit substrate of $\varepsilon_r = 2.2$, $h = 20$ mil. A two-port aperture-coupled microstrip antenna with a transmission loss of 8 dB (due to the radiation of the patch) was located at the path of the oscillator feedback loop. An amplifier [designed using the NEC 32 484A high electron-mobility transistor (HEMT)] with about 10-dB small-signal gain was also placed in the loop to reach the gain requirement for oscillation. In addition, to satisfy the oscillation phase requirement, the electrical length of the feedback loop was adjusted to be a multiple of 360°. The finished active antenna oscillated at the frequency of 9.85 GHz. In this two-element array, part of the oscillating power of the first oscillator was drawn out to the microstrip control line by a −15-dB directional coupler located after the oscillator amplifier, and amplified by an amplifier (with the same design as those in the oscillators) on the control line. The amplified signal was then injected to the second oscillator through another −15-dB coupler located before the oscillator amplifier. The control line was terminated by a 50-Ω load to absorb the rest of the power on the line. Points A and B in the figures are two reference points for considering the coupling phases.

For this system, there exist two kinds of injection-locking signals to the oscillators. One is the mutual coupling $(c_m e^{-j\phi_m})$ through free space, which is a function of the distance between antennas. The other is the signal through the control line $(ke^{-j\phi_c})$, which is unilateral due to the presence of the amplifier. By means of the series oscillator circuit and using the Van der Pol model for the active device, a phase dynamic equation can be derived as follows [4]:

$$\frac{d\theta_i}{dt} = \omega_0 + \frac{\omega_0}{2Q}\,\mathrm{Im}\left(\frac{V_i^{\mathrm{inj}}}{V_i}\right), \qquad i = 1, 2. \tag{1}$$

Here, the two oscillators have been assumed to possess the same free-running characteristics, i.e., the same quality factors Q and free-running signal amplitudes A_0 and frequencies ω_0. $V_i (= A_i e^{j\theta_i})$ is the complex signal in the ith oscillator, with A_i being the oscillation amplitude and θ_i the instantaneous phase. V_i^{inj} is the injection signal to the ith oscillator. Since the injected signals are much smaller than the oscillator signals (i.e., $|V_i^{\mathrm{inj}}| \ll |V_i|$), the oscillation amplitudes would remain near the free-running ones $(A_i \approx A_0)$.

By inserting $V_1^{\mathrm{inj}} = c_m e^{-j\phi_m} V_2$ and $V_2^{\mathrm{inj}} = c_m e^{-j\phi_m} V_1 + ke^{-j\phi_c} V_1$ into (1), and then subtracting the resultant two equations with each other, one obtains

$$\frac{d\Delta\theta}{dt} = \frac{\omega_0}{2Q}\Big[c_m \sin(\phi_m - \Delta\theta) - c_m \cdot \sin(\phi_m + \Delta\theta) - k\sin(\phi_c + \Delta\theta)\Big] \tag{2}$$

where $\Delta\theta (= \theta_2 - \theta_1)$ is the phase difference of the two oscillator signals. When the system becomes steady-state, the time variation $(d\Delta\theta/dt)$ vanishes, from which the phase difference $\Delta\theta$ is solved as follows:

$$\Delta\theta = \tan^{-1}\left(\frac{\mp(k/c_m)\sin\phi_c}{\pm(2\cos\phi_m + (k/c_m)\cos\phi_c)}\right). \tag{3}$$

It is seen that the phase difference $\Delta\theta$ of the oscillator signals is a function of the normalized coupling coefficient k/c_m, the phase ϕ_m of the mutual coupling, and the phase ϕ_c of the control line signal. It is also noticed that there are two possible solutions of $\Delta\theta$ in (3), which have a 180° difference. To determine which solution is stable, one inserts a perturbation term δ to the phase difference $\Delta\theta$ in (2) [4], yielding a differential equation of δ as follows:

$$\frac{d\delta}{dt} = \alpha\delta \tag{4}$$

with

$$\alpha = -\frac{\omega_0}{2Q}\Big[c_m\cos(\phi_m + \Delta\theta) + c_m\cos(\phi_m - \Delta\theta) + k\cos(\phi_c + \Delta\theta)\Big]. \tag{5}$$

When the stability factor α is negative, the perturbation δ will be eventually damped down, corresponding to a stable phase difference $\Delta\theta$.

It is noticed that the mutual-coupling phase ϕ_m can be expressed as $\phi_m = k_o d + \psi$, with k_o being the free-space propagation constant, d the distance between antennas, and ψ a correction term for the consideration of near-field effects [4]. The value of ψ is determined empirically.

The proposed scanning mechanism can be observed from (3). When the amplifier on the control line is turned off, the coupling coefficient k of the control line approaches zero and the phase

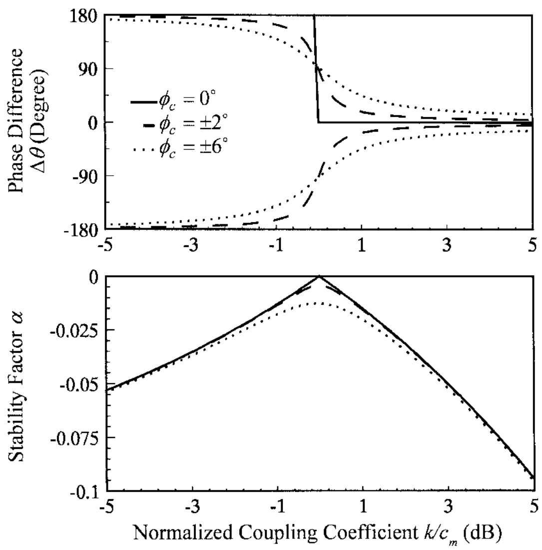

Fig. 2. Variations of the phase difference between antennas and the corresponding stability factor as functions of the normalized coupling coefficient k/c_m. $\phi_m = 120°$.

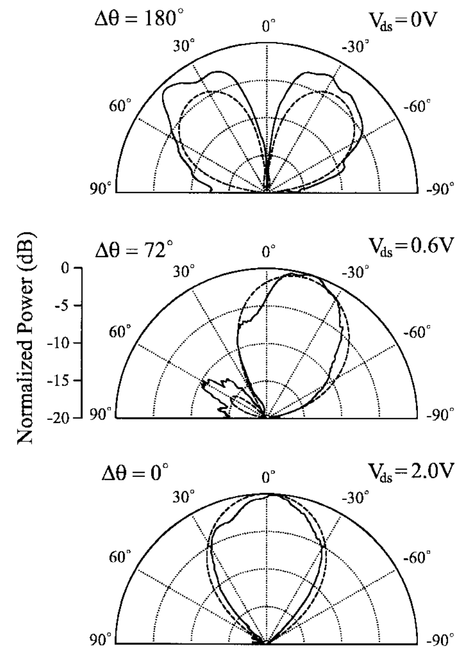

Fig. 3. Measured (solid lines) and calculated (dashed lines) radiation patterns of the two-element active antenna array.

difference $\Delta\theta$ between oscillators is only related to the phase of the mutual coupling. ($\Delta\theta = 0°$ when $\cos\phi_m > 0$, and $\Delta\theta = 180°$ when $\cos\phi_m < 0$). As the bias of the amplifier is increased gradually, the amplifier gain and, thus, the coupling coefficient k increase, resulting in a change of the phase difference $\Delta\theta$. When the amplifier is fully turned on, the coupling coefficient k can be designed to be much larger than the mutual-coupling factor c_m and, from (3), the phase difference $\Delta\theta$ is totally controlled by the phase delay ϕ_c in the control line. Therefore, by suitably biasing the amplifier, the radiation beam of the array can be steered.

III. Results

The correction term ψ for calculating the mutual-coupling phase ϕ_m was estimated by experiments to be $-97°$. The distance d between antennas was chosen as $0.6\lambda_o$, resulting in a mutual-coupling phase ϕ_m of about 120°. The measured mutual-coupling coefficient c_m at this distance was -19 dB. Fig. 2 depicts the calculated variations of the phase difference $\Delta\theta$ and the corresponding stability factor α as functions of the normalized coupling coefficient k/c_m. (Here, only the stable solutions $(\alpha < 0)$ are presented.) The results for $\phi_c = 0°$, $\pm 2°$, and $\pm 6°$ are shown for comparison. When the amplifier on the control line is turned off $(k \approx 0)$, the phase difference between antennas is controlled by the mutual-coupling effect, corresponding to an out-of-phase radiation mode $(\Delta\theta = \pm 180°)$. As k increases (due to the increase of the amplifier's gain on the control line), the phase differences $\Delta\theta$ change monotonically and approach the coupling phases ϕ_c's of the control line [as can be derived from (3)]. For the case of ϕ_c exactly equal to 0°, the mode jumps from the out-of-phase one to the in-phase one $(\Delta\theta = \phi_c = 0°)$ at $k/c_m \approx 0$ dB. This implies that when changing the gain of the amplifier, one can only *switch* the array between these two modes. However, when ϕ_c is designed slightly smaller or larger than 0°, a smooth transition appears between the initial mode $(\Delta\theta = 180°)$ and the final mode $(\Delta\theta = \phi_c)$. The larger $|\phi_c|$ is, the larger the range of k/c_m between transition margins, and more stable (more negative of α) are the modes in the transition region. Thus, with a control line of phase $|\phi_c|$ slightly larger than 0°, one can smoothly steer the array from the out-of-phase mode to the in-phase mode $(\Delta\theta = \phi_c \approx 0°)$ by increasing the gain of the amplifier. Note that during the steering, $\Delta\theta$ decreases from 180° to about 0° for $\phi_c > 0°$, but increases from $-180°$ to about 0° for $\phi_c < 0°$, which means that, for these two cases, $(\phi_c > 0°$ and $\phi_c < 0°)$, the radiation beams go toward the broadside direction oppositely.

In order to steer the arrays, the destination mode, i.e., the radiation mode with the amplifier on the control line fully on, should be in-phase modes $(\Delta\theta = \phi_c \approx 0°)$. To accomplish this, the phase delay from point A (through the control line) to point B (Fig. 1) was designed to be near 0° by adjusting the microstrip-line lengths. Also, from the results in Fig. 2, to effectively scan the array beams, the normalized coupling coefficient k/c_m should at least range from -3 to 3 dB when the control-line amplifier is biased from the off state to the fully on state. The coupling coefficient k is equivalent to the transmission loss from points A to B, including the coupler losses ($= -15$ dB), the gain ($\approx$10 dB) of the fully on amplifier in the second oscillator, and the variable gain of the amplifier on the control line. The variation of this coupling factor as a function of the bias voltage $V_{\rm ds}$ of the control-line amplifier had been measured, which showed that k changed monotonically from -30 to -10 dB when $V_{\rm ds}$ was increased from 0 to 2 V. Thus, when the bias was changed, the normalized coupling coefficient $(k_{\rm dB} - c_{m,\rm dB})$ did cover the effective transition range for scanning. (Note that $c_{m,\rm dB} = -19$ dB.)

The radiation patterns of the finished active arrays were measured with the bias voltage $V_{\rm ds}$ of the amplifier on the control line increased gradually. Fig. 3 presents the results of $V_{\rm ds} = 0$, 0.6, and 2 V. For comparison, the calculated radiation patterns of a two-element passive microstrip antenna array with appropriate phase differences $\Delta\theta$ are also shown. Before the measurement, the free-running frequency of each active antenna was tuned to be 9.775 GHz by adjusting the gate bias $V_{\rm gs}$ of the oscillator amplifier. When $V_{\rm ds} = 0$ V, the free-space mutual coupling dominated the injection effect so that the measured radiation pattern possessed two symmetrical main beams, corresponding to an out-of-phase mode. As $V_{\rm ds}$ was increased, due to the increasing injection signal from the control line, the right main beam smoothly shifted toward the broadside direction while the left beam shrank and shifted away from the array normal, as shown in the pattern for $V_{\rm ds} = 0.6$ V. Finally, the right beam arrived at the broadside direction and the left beam disappeared when the amplifier was fully turned on ($V_{\rm ds} = 2$ V), corresponding to an in-phase mode. During the scanning, the locked frequency of the array had a little change owing to different power ratios of the injection signals from the control line and free space. However, the deviation of this frequency was less than 0.35% over the entire range of the bias voltages.

IV. Conclusions

In this paper, a new phase-shifterless beam-scanning technique for two-element active antenna arrays has been proposed and demonstrated. Two injection-locking signals, i.e., the mutual-coupling signal through free space and the signal on an extra control line, were used in this technique. A theory modified from that in [4] was derived to explain the scanning mechanism. To demonstrate the scanning technique, an active aperture-coupled microstrip antenna array was constructed and measured. An amplifier with variable gains was fabricated on the control line, so that the coupling signal on the line could be changed. The measured results showed that, when the amplifier bias $V_{\rm ds}$ was increased from 0 to 2 V, the radiation pattern was varied smoothly from the out-of-phase mode to the in-phase mode. During the increasing of the bias voltage, the antenna oscillators were locked stably, with the deviation of the locked frequency lower than 0.35%.

References

[1] J. A. Navarro and K. Chang, *Integrated Active Antennas and Spatial Power Combining.* New York: Wiley, 1996.

[2] J. Lin and T. Itoh, "Active integrated antennas," *IEEE Trans. Microwave Theory Tech.*, vol. 42, pp. 2186–2194, Dec. 1994.

[3] Z. Ding and K. Chang, "Modes and their stability of a symmetric two-element coupled negative conductance oscillator driven spatial power combining array," *IEEE Trans. Microwave Theory Tech.*, vol. 44, pp. 1628–1636, Oct. 1996.

[4] R. A. York, "Nonlinear analysis of phase relationships in quasi-optical oscillator array," *IEEE Trans. Microwave Theory Tech.*, vol. 41, pp. 1799–1809, Oct. 1993.

[5] H.-C. Chang, E. S. Shapiro, and R. A. York, "Influence of the oscillator equivalent circuit on the stable modes of a parallel-coupled oscillators," *IEEE Trans. Microwave Theory Tech.*, vol. 45, pp. 1232–1239, Aug. 1997.

[6] J. Birkeland and T. Itoh, "A 16-element quasi-optical FET oscillator power combining array with external injection locking," *IEEE Trans. Microwave Theory Tech.*, vol. 40, pp. 475–481, Mar. 1992.

[7] K. D. Stephan, "Inter-injection-locked oscillators for power combining and phased arrays," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-34, pp. 1017–1025, Oct. 1986.

[8] P. Liao and R. A. York, "A six-element beam-scanning array," *IEEE Microwave Guided Wave Lett.*, vol. 4, pp. 20–22, Jan. 1994.

[9] S. T. Chew, T. K. Tong, M. C. Wu, and T. Itoh, "An active phased array with optical input and beam-scanning capability," *IEEE Microwave Guided Wave Lett.*, vol. 4, pp. 347–349, Oct. 1994.

[10] W.-J. Tseng and S.-J. Chung, "Analysis and application of a two-port aperture-coupled microstrip antenna," *IEEE Trans. Microwave Theory Tech.*, vol. 46, pp. 530–535, May 1998.

Variational Analysis of a Gap in the Central Conductor of a Rectangular Coaxial Line

Sujit Chattopadhyay and Pradip Kumar Saha

***Abstract*— Rectangular coaxial line (RCL) discontinuity in the form of a gap in the central conductor and steps in the outer conductor in the planes of discontinuity has been analyzed by a variational method. The orthogonal-mode functions of an RCL required for numerical computation of the discontinuity parameters have been determined using the Ritz–Galerkin technique. The capacitances of the equivalent Pi-network are presented as function of gapwidth, normalized frequency, and ratio of outer conductor dimensions for complete characterization of the discontinuity. The low-frequency series and shunt capacitance values are verified against the static capacitances computed separately by the finite-difference technique.**

***Index Terms*— Discontinuity, variational method.**

I. Introduction

The rectangular coaxial line (RCL) has been a subject of investigation, mainly for its TEM characteristic impedance [1]–[4] and also for the TE- and TM-mode spectra [5]. However, only within recent years have the RCL and its square version found applications in microwave circuits, particularly the beam-forming networks in the lower microwave bands. High Q-values and power-handling capability, significant reduction in size (in comparison with the rectangular waveguide because of operation in the TEM mode), high-density packing of transmission lines (due to its geometry), and integration of an entire network on the same block have been noted as the major attributes of the RCL [6].

For realization of various passive components in the RCL, discontinuities and their characterization are the basic ingredients. However, such data are scarce within the literature. The major source of information available appears to be those reported by Sorrentino and others, who have carried out both theoretical and experimental investigation [7]–[9]. Some additional discontinuity-related information may be found in [10] and [11].

This paper's authors have taken up the RCL discontinuity of gap in the central conductor and determined its equivalent circuit using a standard variational technique applicable to symmetric double discontinuity [12]. The structure is made more general by introducing a step in the outer conductor of the RCL at both the planes of discontinuity in the central conductor, such that the dimensions of the rectangular waveguide formed by the gap region are larger than or equal to those of the RCL. In the analysis to obtain the upper bound (UB) for the discontinuity capacitance, it is necessary to expand the aperture electric field in the discontinuity plane in terms of the orthogonal-mode functions of the RCL region. Although Gruner's work [5] is available to calculate the TE–TM eigenvalues of the RCL, complete analysis and computation had to be carried out independently to determine the orthonormal-mode functions of the RCL, which are necessary for the discontinuity problem. Thus, this work is divided into two parts. In the first part, the Ritz–Galerkin technique is applied to generate the normalized fields of the dominant

Manuscript received July 28, 1997; revised June 20, 1998.

S. Chattopadhyay is with the Department of Telecommunication, Government of India, Calcutta 700 0073, India.

P. K. Saha is with the Institute of Radio Physics and Electronics, University of Calcutta, Calcutta 700 009, India.

Publisher Item Identifier S 0018-9480(99)01139-4.