Suspended Resonators for Filters-Reduced λ_g Excitation of Evanescent Cavities Using High Dielectric Constant Feedlines

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***Abstract*—A structure consisting of dielectric-loaded feed lines (such as surface-wave lines similar to Goubau lines) and below-cutoff air-filled cavities can be used to form essentially *L–C* sections. The capacitance is due to electric-field coupling from the feed-line dielectric medium into the below-cutoff section. The inductance results from combining the inductors in the inductive tee equivalent circuit for such below-cutoff sections. Dielectric loading is used to shorten the guide wavelength at the input to the evanescent section, increasing the effective input inductance. The dielectrically loaded feed lines can comprise microstrip, coplanar waveguide (CPW), coplanar stripline (CPS), Goubau lines (surface-wave structures), waveguide, etc. The resulting resonant elements are usable at frequencies below 1 GHz, with small dimensions. If connected to the common ground plane, these L–C sections act as a transmission zero. If "floated," i.e., connected in the "hot" line rather than to the ground plane, the sections form bandpass circuits (transmission poles). The air-filled below-cutoff sections (evanescent mode) are placed in a supporting low dielectric-constant medium (air, Teflon, or similar) with the open end in proximity to the dielectric portion of the feed line and are, thus, termed "suspended." The individual L–C sections can be coupled together using microstrip, surface-wave line, CPW, CPS, finline, waveguide, or lumped elements. Such combinations can be chosen to implement Chebychev, Butterworth, quasi-elliptic, etc. responses. These applications will be covered at a later time.**

***Index Terms*—Bandpass, bandstop, below cutoff, dielectrically loaded, evanacent, Goubau line, resonators, surface wave, suspended line.**

I. Introduction

In an interesting paper [1], Papapolymerou *et al.* present the possibility of exciting a resonant cavity micromachined into a silicon substrate. This cavity supports a dominant-mode resonance for the particular cross section and length employed. In [1], the operating frequencies were in the X-band. In this paper, we present the possibility of exciting mechanically similar cavities, but at much lower frequencies. The cavities will operate as evanescent-mode inductors, and will be shown to resonate when combined with capacitance effectively resulting from electric-field coupling between the open end of the evanescent section and the high dielectric-constant material forming a portion of the lines feeding the evanescent section. Particular examples will be presented for operation as low as 1000 MHz. The resonance effect results from what will be called the "equivalent frequency" principle, by which it is recognized that a below-cutoff section is below cutoff, not to a given frequency, but to the wavelength of energy incident upon it. Table I, (4-1) illustrates that the reactance of the below cutoff section is dependent on the ratio of the wavelength of the incident energy (λ_g) to the cutoff wavelength for the section (λ_c). Thus, shortening that incident wavelength through the use of dielectric loading enables the below cutoff section to be effectively closer to cutoff and, thus, more easily excited. The tee-equivalent series inductance is increased [see Table I, (4-1)], enabling resonance with a smaller capacitor for a particular resonant frequency. The basic structure and principle of the suspended resonators are illustrated in Figs. 1–4. The defining equations are presented in Table I. In this paper, the below-cutoff sections are round, but any closed

TABLE I
EQUIVALENT-CIRCUIT COMPONENT VALUES FOR FIG. 4

$$X_e = Z_o \tanh\left(\frac{\gamma \ell}{2}\right) + \frac{Z_o^2 \tanh\left(\frac{\gamma \ell}{2}\right)}{Z_o \tanh\left(\frac{\gamma \ell}{2}\right) + \frac{Z_o}{\sinh(\gamma \ell)}} \tag{4-1}$$

Z_o (for round cross section sector with cut-off wave length of λ_c)

$$Z_o = \frac{377}{\sqrt{\left(\frac{\lambda_g}{\lambda_c}\right)^2 - 1}} \tag{4-2}$$

$$\gamma = \left(\frac{6.28}{\lambda_g}\right)\sqrt{\left(\frac{\lambda_g}{\lambda_c}\right)^2 - 1} \tag{4-3}$$

The values of Zo & from [2], and guide wavelength from the dielectric constant in the surface wave feed lines. Derivation of C is presented in the Appendix.

$$C = \frac{2\pi \varepsilon_r \varepsilon_o r \sqrt{4d^2 + r^2}}{4\sqrt{d^2 + r^2} - r} \tag{4-4}$$

r = radius of cylinder, d = thickness of dielectric layer in surface wave line structure, C is effective total circuit static capacitance.

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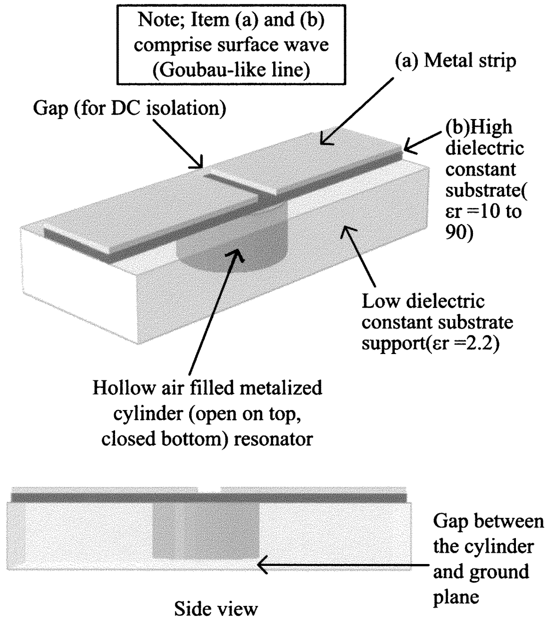


Fig. 1. Evanescent suspended bandpass resonator (series transmission pole).

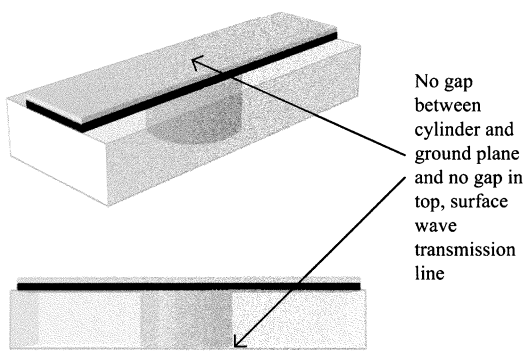


Fig. 2. Evanescent suspended bandstop resonator (shunt transmission zero).

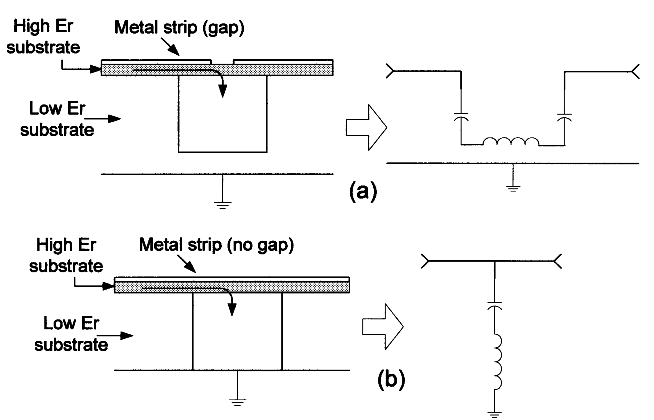


Fig. 3. Equivalent-circuit elements for both: (a) bandpass and (b) bandstop cases; values of capacitance C given from Table I, (4–4).

shape would work (e.g., rectangular, elliptical, etc.,open on at least one end). The equivalent circuits are also shown in these figures, with the values of the elements shown. Inductances stem from the single-mode tee equivalent circuit of [2] and are presented in Figs. 3 and 4 [as shown in Table I, (4-1)–(4-3)], with the capacitance value given in Table I, (4-4), derived from image

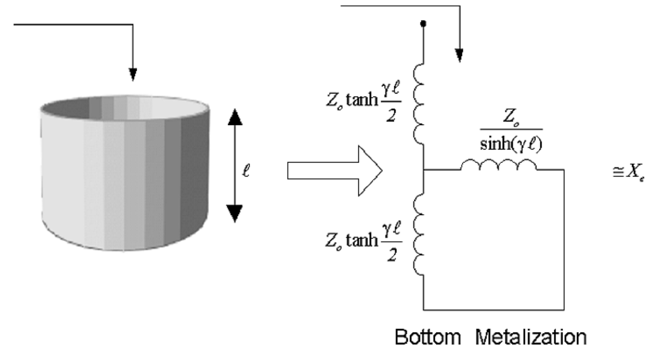


Fig. 4. (a) Metallized wall and bottom below-cutoff cross section. (b) For a single-mode below cutoff, the equivalent circuit is a short-circuited tee.

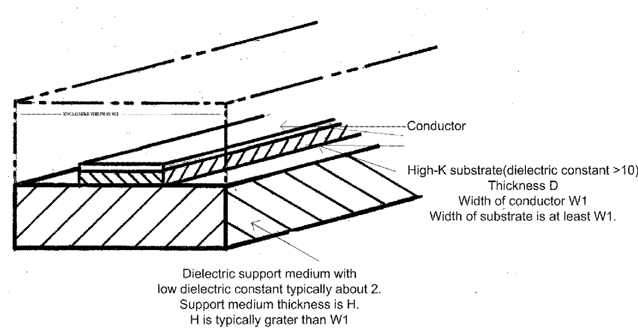


Fig. 5. Surface-wave line configuration. Enclosure width is W_2, line and dielectric widths are W_1, high dielectric-constant substrate thickness is d ($\varepsilon r > 10$), support thickness is H, and $\varepsilon r = 2$. For surface wave, $H > W_1$.

theory and contained in the Appendix . The closed cup suspended above the ground plane provides a series resonant circuit and, thus, a transmission pole, while the cup directly connected to the ground plane implements a transmission zero. The combinations of series and shunt resonant circuits can be used to implement the usual variety of bandpass and bandstop filter circuits, with the shunt circuit used for additional transmission zeros, as required by the desired topology.

II. Excitation and Interconnection of Resonators

The feed lines used for exciting the evanescent resonators are termed "Goubau like" in honor of G. Goubau who first described the phenomenon whereby RF energy propagates close to an isolated dielectric-coated conductor [3]. In this paper, we use a microstrip-like configuration, but with the ground plane far from the isolated conductor. This conductor is supported on a high dielectric-constant substrate and, thus, most of the energy is bound by the dielectric, with the wavelength set by the dielectric constant of the substrate. The relatively high dielectric constant (>10) essentially eliminates radiative losses from the line and, thus, ensures low-loss transmission of energy. The configuration is illustrated in Figs. 1, 2 and 5, with typical impedance data in Fig. 6. It can be seen that, as the distance of the line to the ground plane decreases, the line approaches microstrip. However, as the line moves away from the bottom, the impedance is primarily a function of the ratio of the enclosure width W_2 to the line/dielectric width W_1, and energy is essentially bound by the conductor and retained in the dielectric layer. It has been

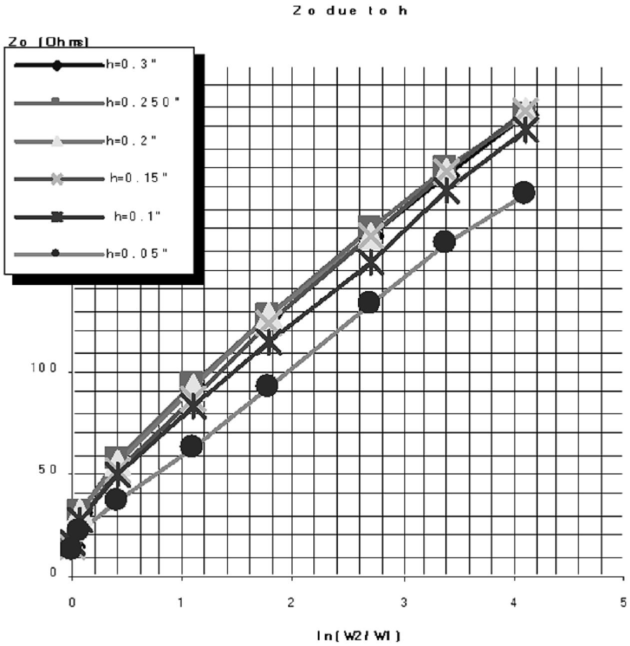


Fig. 6. Z_0 versus $\ln(W_2/W_1)$ for various values of H.

found that the line Z_0 displays essentially the same dependence on H for a wide range of W_2 and is, thus, primarily a function of the ratio W_2/W_1 for $H > W_1$. Thus, Fig. 6 can be used in the design of interconnecting lines for implementing various filter topologies, as well as for excitation of the resonators. It has been found that, as long as the guide wavelength has been reduced in the immediate vicinity of the below cutoff resonator (with a short length of high dielectric-constant surface-wave line), the majority of the interconnecting lengths of the transmission line can be approximated with a lumped low-pass network. This equivalent network is required to provide the same input impedance and phase shift as the transmission line that resulted from the original synthesis. This lumped equivalent network has another significant advantage: it does not display a periodic response and, thus, the stopband of the bandpass or bandstop structure also does not display periodicity.

III. Resonators

As shown in Figs. 1 and 2, the resonators can be connected to implement either series or shunt resonant equivalent circuits. The effective inductance of any below cutoff section (round in this paper, but also possibly rectangular, elliptical; any cross section displaying dispersion and a high-pass cutoff characteristic) results from the fact that the cutoff wavelength for the section is shorter than the incident signal wavelength. With this new approach, high dielectric-constant loading is used to modify the incident signal wavelength, reducing the difference between the cutoff wavelength and incident signal wavelength. Without the reduction in signal wavelength resulting from dielectric loading, the effective series inductance in the equivalent circuit would be lower and more resonating capacitance would be required, in either the series or shunt case, for a particular resonant frequency. The equivalent series inductance is proportional to the square root of the substrate dielectric constant. Simulated and

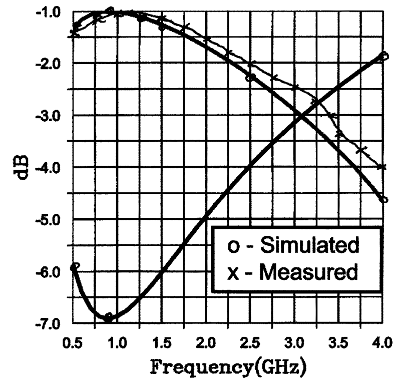


Fig. 7. Bandpass case: $F_0 = 1.03$ GHz (measured), 0.93 GHz (simulated), dielectric constant $= 25$ in simulation, 20 actual, substrate thickness is 0.004 in.

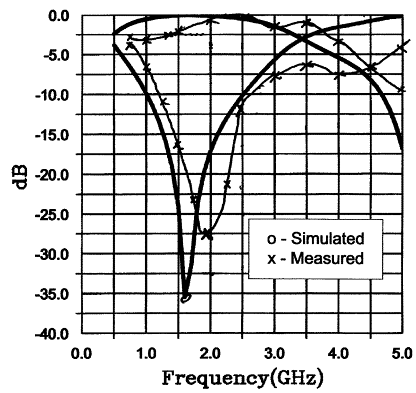


Fig. 8. Bandstop case: $F_0 = 1.82$ GHz (measured), 1.65 GHz (simulated), dielectric constant $= 25$ in simulation, 20 actual, 0.004 in thick.

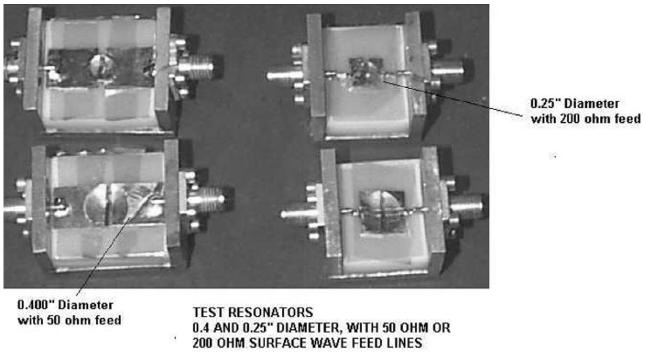
measured data are shown for a typical series and shunt case in Figs. 7 and 8. In the frequency range between 1–2 GHz, the effective unloaded Q for these resonators is approximately 400. The loss tangent for the dielectric material used is measured as 0.002, thus, the majority of the resonator loss is dissipation in the dielectric. A physical comparison of the transmission line and lumped interconnects is shown in Fig. 9.

IV. Operational Comments

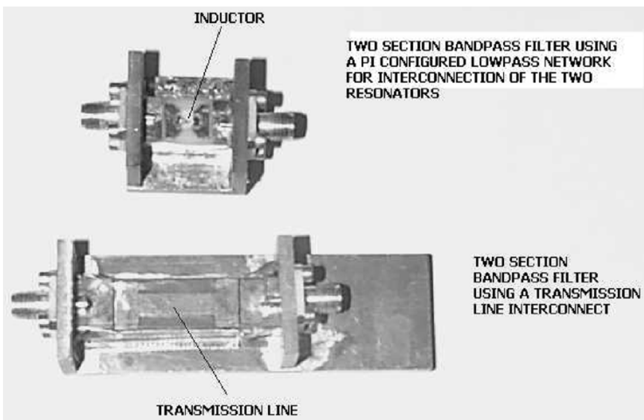
Thus, the dielectric loading has the effect of allowing resonance at lower frequencies without using large resonating capacitors. Further, the dielectric loading does not sacrifice a major advantage of evanescent resonant structures, i.e., very wide stopbands, because spurious passbands do not occur until frequencies exceed the cutoff frequency for the below-cutoff section. The cutoff frequency of the below cutoff section is **NOT** affected by the dielectric loading of the feedlines. At some point, it is intended to use ferroelectric dielectrics for implementation of electrically tunable resonators.

V. Filters and Conclusions

A large number of filters have been built, including two to six-pole designs, both ladder and cross-coupled topologies, from approximately 400 MHz to 4 GHz. It is believed that the approach is applicable down to 100 MHz. Given the technology to fabricate very small closed surfaces, the approach could also



(a)



(b)

Fig. 9. (a) Bandpass and bandstop resonators. (b) Two resonator bandpass types using either transmission line or lumped inverters.

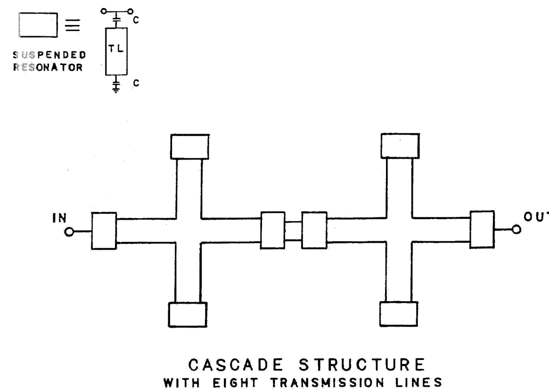


Fig. 10. One proposed multiresonator connection.

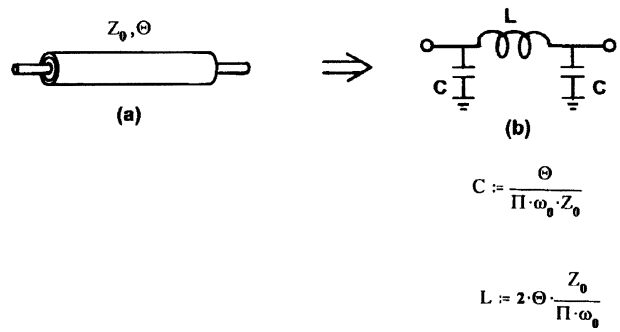


Fig. 11. Transformation of a series transmission line shown in (a) into a low-pass pi-equivalent shown in (b). θ is in radians, ω_0 is filter center frequency in radians. Final values are adjusted via optimization.

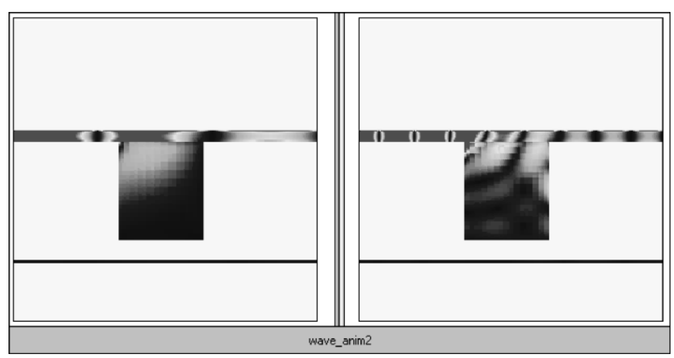


Fig. 12. Propagating wave comparison. Equivalent frequency principle: higher K shortens wavelength and has same effect as higher frequency with low K increases reactance of below cutoff resonator.

be applied to "on-a-chip" filter configurations from 100 MHz to at least 10 GHz. In some cases, the individual resonators have been connected with lengths of the surface-wave line, with each length representing an inverter, as required by basic synthesis. In more recent implementations, the interconnecting transmission lines have been replaced with lumped equivalents, as described above. Many topologies are possible, including cross-coupling of the bandpass resonators, individual transmission zeros placed with the bandstop resonators, etc. Two resonator networks are shown in Fig. 9 (one with a transmission line and one with a lumped interconnect) with one possible higher order interconnection in Fig. 10. The low-pass equivalent technqiue is describe in [4] with the transformation equations summarized in Fig. 11. The wavelength-shortening effect is shown in Fig. 12. The filters will be fully described in a future paper.

Appendix
Lower Order Capacitance Term Derivation

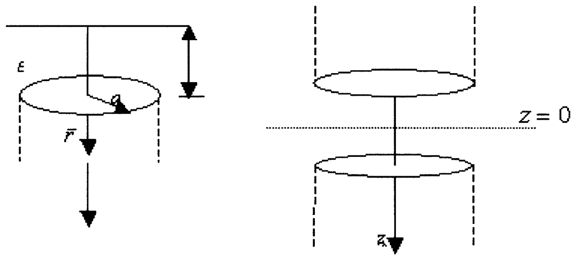
The cylindrical resonator of radius a and height l is made of an electrical conductor. The top is covered by a dielectric layer with permittivity $\varepsilon = \varepsilon_o \varepsilon_r$ and thickness d, where $d \ll a$. The metal strip line on top of this layer is used to excite a surface wave. Generally, the width of the strip line exceeds the diameter of the resonator when impedance matching is considered.

In determining the lowest order term for the capacitance between the strip line and resonator, the following assumptions are made.

- The finite width of the strip line is assumed to be infinite.
- The induced charge on the cylinder is confined only to the rim due to its metallic nature.

The image theory is used based on the above assumptions and the equivalent image diagram is obtained. Fig. 13(b) is based on the original geometry of Fig. 13(a).

In order to solve the image geometry in Fig. 13(b), consider a single circular filament of charge density ρ_l located on the xy-plane, as shown in Fig. 14.



(a) (b)

Fig. 13. Surface-wave excitation of a cylinder resonator. (a) Original geometry. (b) Equivalent image.

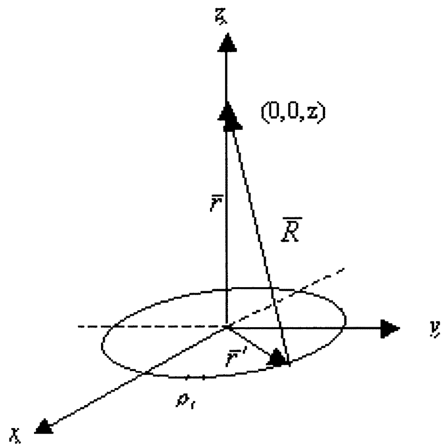


Fig. 14. Circular line charge density.

The position vectors are $\bar{r} = z\hat{z}$ and $\bar{r}' = a\hat{\rho}$. The corresponding distance is given as

$$R = |\bar{r} - \bar{r}'| = \left(z^2 + a^2\right)^{1/2}. \tag{A-1}$$

The differential electric field is given as

$$d\bar{E}\,(\bar{r}) = \frac{1}{4\pi\varepsilon}\frac{\rho_l\,(\bar{r}')\,adl'\,(\bar{r} - \bar{r}')}{|\bar{r} - \bar{r}'|^3} \tag{A-2}$$

and the total field along the $+z$-axis (any other observation point off the axis will require formulation in terms of elliptical functions) is

$$\bar{E}\,(0,0,z) = \frac{1}{4\pi\varepsilon}\int_0^{2\pi}\frac{\rho_l a d\phi'\,(z\hat{z} - a\hat{\rho})}{(z^2 + a^2)^{3/2}}. \tag{A-3}$$

Note that $dl' = ad\phi'$.

The evaluation of the above integral yields only a z-component of the E-field along the $+z$-axis as

$$E_z\,(0,0,z) = \frac{az}{2\rho}\rho_l\frac{1}{(z^2 + a^2)^{3/2}}. \tag{A-4}$$

Field components other than the z-component vanish due to symmetry. Using the above result in Fig. 13(b) for the equivalent image yields for the E-field between the rings as

$$E_z = \frac{-a(d - z)\rho_l}{2\varepsilon\left[(d - z)^2 + a^2\right]^{3/2}} - \frac{a(z + d)\rho_l}{2\varepsilon\left[(d + z)^2 + a^2\right]^{3/2}} \tag{A-5}$$

and the resulting potential difference between the two rings can be obtained as

$$V_0 = -\int_{-d}^{d}\bar{E}\cdot d\bar{l} = \frac{a\rho_l}{\varepsilon}\left[\frac{1}{a} - \frac{1}{(4d^2 + a^2)^{1/2}}\right]. \tag{A-6}$$

Since the total charge on any ring is

$$|Q| = \rho_l 2\pi a.$$

The equivalent capacitance is then

$$C = \frac{|Q|}{V_o} = \frac{2\pi\varepsilon a\left[4d^2 + a^2\right]^{1/2}}{(4d^2 + a^2)^{1/2} - a}. \tag{A-7}$$

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