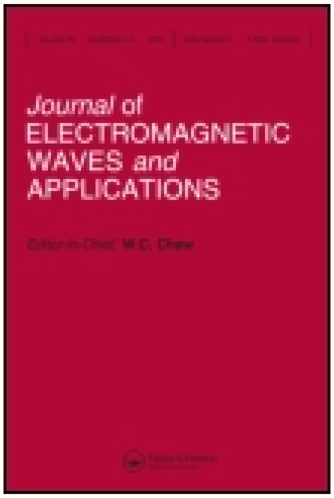

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ANALYSIS OF AN ELECTROMAGNETICALLY COUPLED MICROSTRIP RING ANTENNA USING AN EXTENDED FEEDLINE

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Abstract—A new method for the computation of input impedance of an annular ring microstrip antenna, fed electromagnetically by a microstrip line, and operating in its fundamental mode is presented. The method based on cavity model in conjunction with circuit theory is compared with simulation (IE3D$^{\text{TM}}$) and measurement based results. It is shown that this method of computation leads to accurate prediction of impedance for such structures with minimum of computational complexity, thereby making it an attractive tool for accurate and quick design estimate for many applications.

1. INTRODUCTION

Narrow bandwidth has been a concern for the optimum use of microstrip antennas. Electromagnetically coupled feeding of the antennas is one of the methods [1] to enhance the bandwidth of the antenna. It is known that a microstrip antenna built on electrically thick substrate will exhibit large impedance bandwidth. However, for thicker substrates, probe feeding remains a detrimental issue due to larger component of probe inductance. On the other hand, the electromagnetic coupling method, where the antenna is excited by a feed line below the antenna is an attractive mean of feeding a

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microstrip antenna on thick and multi-layer substrate. Different mathematical methods were used to analyse the electromagnetically coupled feed for dipole antennas [2, 3]. The ring antenna has been rigorously analysed using Galerkin's method, the method of moments and circuit analysis [4]. Annular ring antenna with stacked circular disc is presented in [5]. In this case, the given structure, where the annular ring is probe fed and is coupled electromagnetically to the circular disc on top, is analyzed using modal expansion cavity model. The theoretical results indicate nearly 14bandwidth. Such interesting propositions for electromagnetically coupled (also known as proximity coupled) patch radiators continue to be placed before the scientific community. Neural network based optimization and characterization of electromagnetically coupled patch antennas have been reported [6]. In one recent reference [7], authors have reported the design and performance of circularly polarized loaded square patch fed through proximity coupling. The antenna is developed using simulation based methods. Multi-layer feeding for broadband reconfigurable microstrip antenna is displayed in [8], where the antenna again is designed and optimized using full-wave simulation. The ability of electromagnetic coupling to increase bandwidth is aptly demonstrated in [9], where the authors have designed a dual circularly polarized antenna using L-strip. The cited references demonstrate that electromagnetic coupling method continues to be used for developing innovative antenna structures for commercial markets. However, these structures are developed based on accurate but cost-intensive full-wave simulators. Even though a full wave analysis of such structure yields very accurate results, considerations such as quick design estimates make closed form expressions highly desirable for practicing engineers. Quick design estimates require analytic expressions to be solved for given constraints and objectives. The continuing need for such techniques can be judged by the fact that scientists keep attempting to obtain equivalent circuit parameters of conventional microstrip antennas. As for example, in [10] the authors have attempted to compute the wideband lumped equivalent model of a rectangular antenna considering the first few radiating resonances. It is shown that such methods are very useful for design estimates.

This paper proposes a new method of analysis, based on equivalent circuit concepts, for an electromagnetically coupled ring antenna yielding reasonably accurate results. The novelty of the proposed method can be judged from the following fact. The proposed computation method translates a well established cavity model for probe fed ring antenna, similar to one given in [11], to an equivalent circuit when the feeding method is changed to proximity coupling (instead of coaxial probe feeding). Again, the translation is done using standard circuit model and transmission line concepts. It is shown that an application of such fundamental concepts to the structure leads to accurate prediction of the input impedance. This method is generic enough to be extended to other patch structures. The proposed method is simple to execute and accurate enough to be used as a first hand design estimate before a computationally intensive and costly full-wave solver is invoked to fine tune the end design. To the best of authors knowledge, such technique has not been attempted before on electromagnetically coupled microstrip radiators fed by a transmission line.

2. THEORY

The geometry of an electromagnetically coupled ring antenna is shown in Fig. 1. The ring antenna is constructed on two layer substrate of thickness h_1 and h_2 with relative dielectric constants ε_1 and ε_2 respectively. The inner radius of the ring is 'a' and the outer radius is 'b'. The ring is fed underneath by a transmission line of length 'l' and width 'w' with an insertion of 's' from the outer edge of the ring. The equivalent circuit of the configuration is given in Fig. 2. Using the equivalent circuit, the computation of the impedance seen by the coaxial probe at the end of the feeding transmission line is done in three steps. For this unique translation method, in the first step, the impedance of a probe-fed ring is computed for a virtual probe location at $r_0 = b - s$ with respect to the centre of the ring. It is then transformed by a capacitive tap constituting the overlap

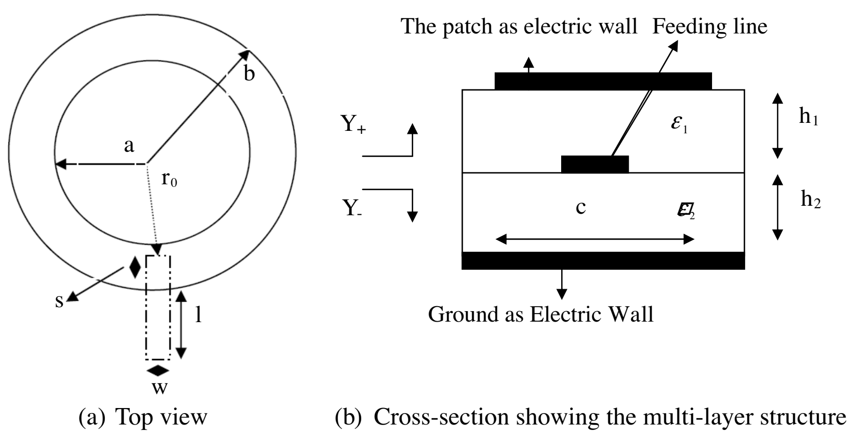

Figure 1. Geometry of the microstrip line fed ring resonator.

area of the ring and the microstrip line. In the final step, the net impedance is transformed at the line end using a classical transmission line transformation. For the first step, the impedance of a probe fed ring resonator at a virtual probe location r_0 is given by [11]

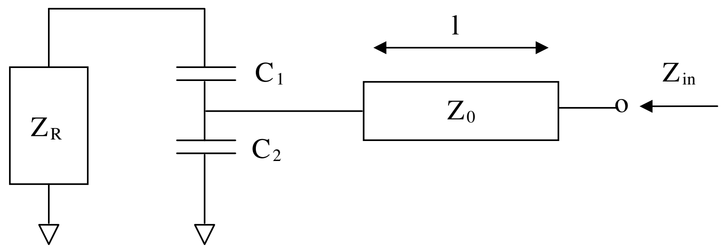

Figure 2. Equivalent circuit of the microstrip line fed ring resonator.

$$Z_R = \mu^2 h^2 \times \sum_n \sum_p \frac{(j\omega + A) C_n^{(1)^2} \omega^2 \left[\left(J_n(k \cdot r_0) Y_n'(kb_e) - J_n'(kb_e) N_n(k \cdot r_0) \right) J_0\left(\frac{nd_f}{2r_0}\right) \right]^2}{(j\omega - C)(j\omega + A) + \omega_{np}^2} \tag{1}$$

The parameters A & C are given as

$$A = Z_s \int |H_i|^2 dS \quad \text{and} \quad C = -Y_w \int |E_i|^2 dS \tag{2}$$

where ω_{np} corresponds to the resonance of the eigen-mode corresponding to the TM_{np} mode, 'b_e' and 'a_e' represent the effective radius due to fringing fields, $J_n(kr)$ is the Bessel function of order n, and $J_n'(kr)$ is the derivative of first order Bessel function. For the computation of impedance, the expression for the wall admittances and mutual admittances has been obtained from [12].

In the second step, the net capacitance on the line underneath the ring is evaluated. The net capacitance seen by the transmission line with respect to the two conducting planes on top and below is evaluated using the accurate method outlined in [13]. The expressions required to evaluate the capacitance are given in the subsequent section.

Let the transmission line of characteristic impedance Z_0 and length l be loaded by Z_R, the impedance of ring and the capacitive tap, then the combined load on the transmission line may be denoted as Z_L. Therefore the input impedance seen by a source is given as

$$Z_{in} = Z_0 \frac{Z_L + jZ_0 \tan(\beta l)}{Z_0 + jZ_L \tan(\beta l)} \tag{3}$$

For our analysis we have considered a lossless line. Therefore by using Equations (1), (2) and the net capacitance figure we can evaluate the input impedance of the ring resonator. An exact estimation of the capacitance in the multilayer structure is done using variational analysis in conjunction with transverse transmission line theory and detailed in [13] For sake of completion, the expressions for capacitance evaluation are listed in the following section.

3. ANALYTICAL METHOD OF CAPACITANCE COMPUTATION

Following [13], it is seen that capacitance formulation has two components namely parallel plate capacitance between the two layers and a fringe capacitance along all the open ends of the transmission line. Fig. 1(b) displays the structure under study. For this structure the admittance are computed on the charge plane (line) as Y_+ and Y_- Substituting these parameters in the following expression gives the capacitance between the line and the respective walls:

$$C = \frac{(1 + 0.25A)^2}{\sum\limits_{n\ odd} (T_n P_n / Y)}, \tag{4}$$

where the admittance on the charge plane defined by Y is given as

$$Y = Y_+ + Y_-, \tag{5}$$

$$Y_+ = \varepsilon_0 \varepsilon_1 \coth \beta h_1, \tag{6}$$

$$Y_- = \varepsilon_0 \varepsilon_2 \coth \beta h_2. \tag{7}$$

The expressions for other parameters in Equation (3) are given in [13] and they need to be used by considering the transverse wall to wall separation as large ($c \gg 5w$).

4. RESULTS

The theory formulated in the previous section is validated by comparing the computed results with simulation based results (IE3D from M/s Zeland Softwares). The dimension of the ring is chosen as follows: $a = 30\,\mathrm{mm}$, $b = 60\,\mathrm{mm}$, $h_1 = h_2 = 0.787\,\mathrm{mm}$ and $\varepsilon_1 = \varepsilon_2 = 2.2$. A microstrip line of given width w and length l is drawn on height $h_2 = 0.787\,\mathrm{mm}$ with an insertion of s underneath the ring. The characteristic impedance of the line in the proposed environment is evaluated and used in (2). If the proposed insertion is,

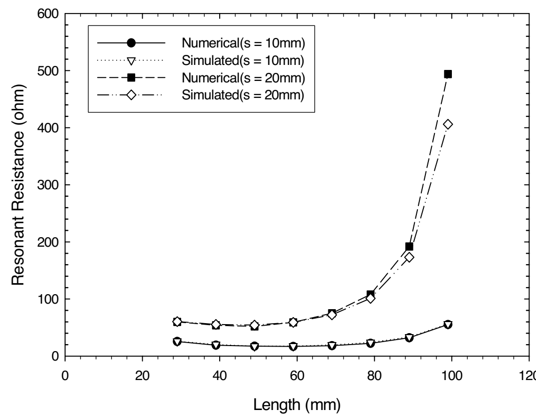

Figure 3. Comparison of present theory with simulated data for $w = 0.5\,\mathrm{mm}$ for TM_{11} mode (peak value resonant resistance).

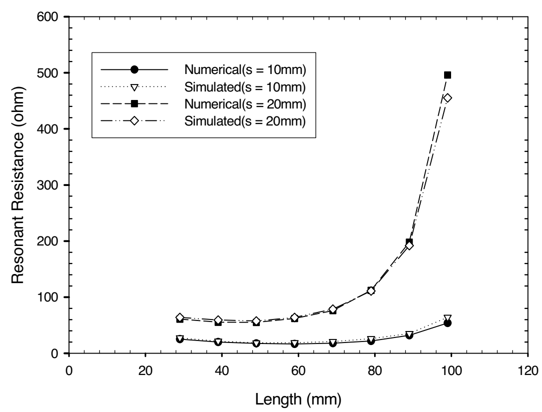

Figure 4. Comparison of present theory with simulated data for $w = 0.95\,\mathrm{mm}$ for TM_{11} mode (peak value of resonant resistance).

for example, 10 mm underneath the ring (from the outer edge), then (1) is evaluated for the input impedance of the ring considering a probe location of $r_0 = 50\,\mathrm{mm}$. For comparison, two different widths of the line are chosen with two different insertion namely $s = 10\,\mathrm{mm}$ and $s = 20\,\mathrm{mm}$.

From Figs. 3 and 4, it is seen that excellent agreement exists in the resonant resistance of TM_{11} mode of the loaded ring resonator between theory and simulation with transmission line loading. The deviation begins to develop when the width of the line is increased.

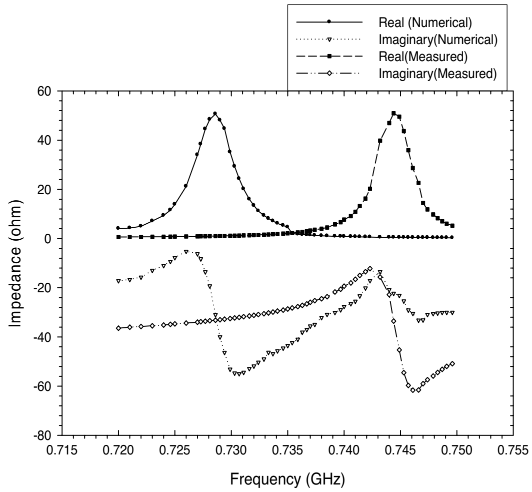

Figure 5. Input impedance of fundamental mode of ring resonator loaded with transmission line ($a = 30\,\mathrm{mm}$, $b = 60\,\mathrm{mm}$, $w = 0.95\,\mathrm{mm}$, $l = 29\,\mathrm{mm}$).

The estimation error in resonant frequency is approximately 2% for dominant mode and 1.5% for higher mode. The computed and measured values of the input impedance for TM_{11} mode are shown in Fig. 5. It is seen that there is a very good agreement between the computed values and the measured results, except for a 2.5% error in the prediction of the resonant frequency.

5. CONCLUSION

The proposed theory and its good agreement with numerical and measured results show that such simple steps can be used to make accurate design estimates for such structures even exploring the possibility of using ring in its fundamental mode thereby reducing the size.

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