A USEFUL EQUIVALENCE FOR THE INPUT REACTANCE SEEN BY THE COAXIAL LINE IN A BROADWALL COAXIAL-MICROSTRIP LAUNCHER

Chinmoy Das Gupta and R.S. Tomar

Department of Electrical Engineering
Indian Institute of Technology
KANPUR (India)

Abstract

The broadwall coaxial-microstrip launcher has been studied from the point of view of modelling the reactance seen at the coaxial port. It turns out that, for the purpose of writing the input reactance, the feeding probe can be treated as a thin cylindrical antenna situated in an homogeneous infinite space.

Summary of the Paper

The use of coaxial probes to feed microstrips and striplines is quite well-established. The most commonly employed techniques consists in launching the electromagnetic energy from the edge of the line (called 'end launcher').There, however, are situations in which it becomes convenient to launch into the microstrip/stripline through the ground plane rather than from the edge. Such a configuration, called a broadwall launcher, is shown, in its microstrip form, in Fig. 1. In this configuration, the coaxial line is mounted on the ground plane and the central conductor of the coaxial line penetrates the substrate through a cylindrical hole to feed the strip.

The broadwall launcher is particularly useful when small ground plane spacings are involved. Although some discussion on this kind of launchers is available in literature [1], directly usable results regarding the analysis/ synthesis of such launchers seem to be generally unreported. The intent of the present paper is to report an investigation on the input reactance seen by the coaxial line in such a launcher. The equivalence established herein will help in quantitative modelling of such launchers.

The basic idea used in the investigation is quite simple. We let the free-end of the coaxial line be terminated into a movable short-circuit, as shown in Fig. 2. The transmission coefficient

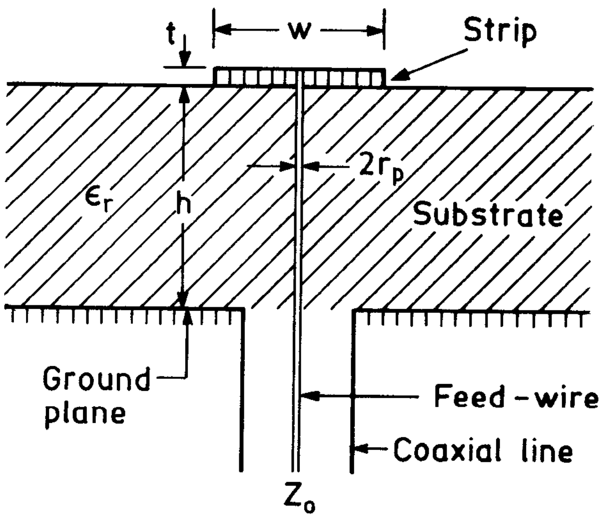

Fig. 1: Broadwall coaxial-microstrip launcher.

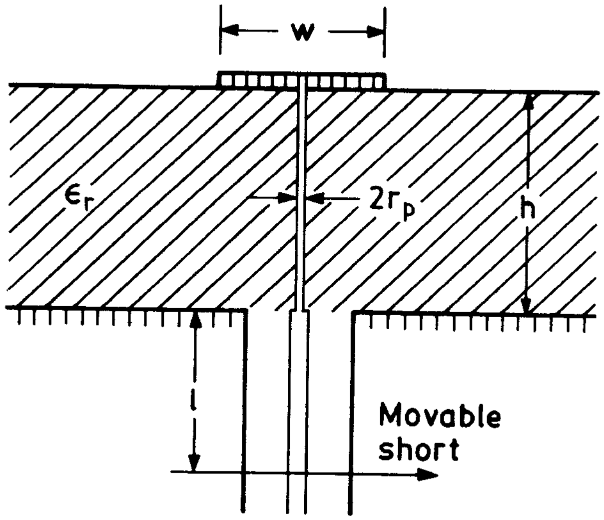

Fig. 2: Proposed resonant structure.

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S_{12} between the microstrip ports will then exihibit a sharp minimum in its variation with the position of the movable short, as shown qualitatively in Fig. 3*. In fact, an accurate closed - form expression for S_{12} can easily be

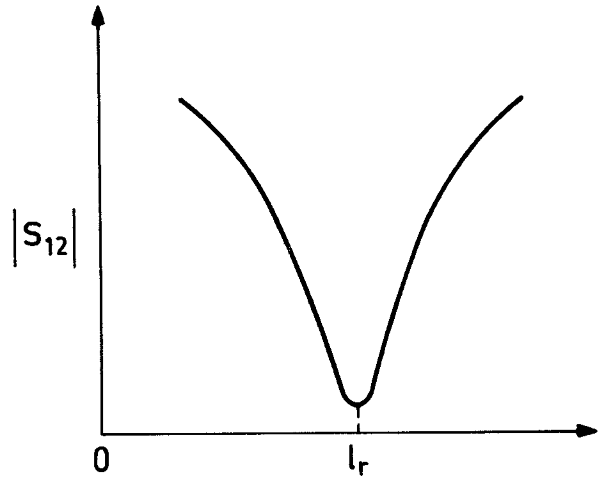

Fig. 3: Variation of transmission-coefficient with the position of the movable short.

derived by first converting the resonant structure of Fig. 2 into an equivalent planar waveguide configuration (for details on planar waveguide equivalence of a microstrip, see [3]) as shown in Fig. 4, and then by analysing this configuration along lines identical to those followed in [2]. The condition for minimum transmission between the microstrip ports, or the resonance condition, can then easily be derived and the equivalent circuit at resonance be shown to be as given in Fig. 5. Subsequent

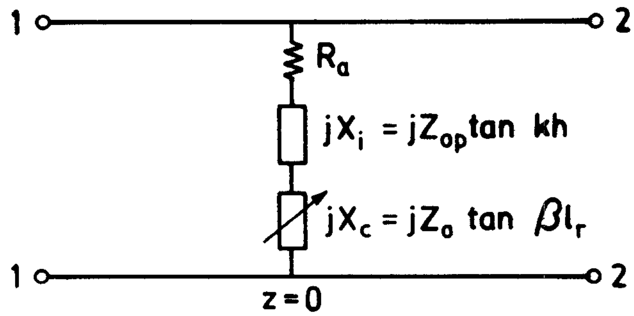

Fig. 5: Near-resonance equivalent circuit.

experimental verification of this resonance condition (see Tables I and II) reveals that the prove characteristic impedance Z_{op} should be chosen to be given by

$$Z_{op} = \frac{120}{\sqrt{[\varepsilon_{re}(f)]}}\,[\ln \frac{h}{r_p} - 1]$$

in order to get the best possible agreement between theoretical and experimental values of the resonant length.

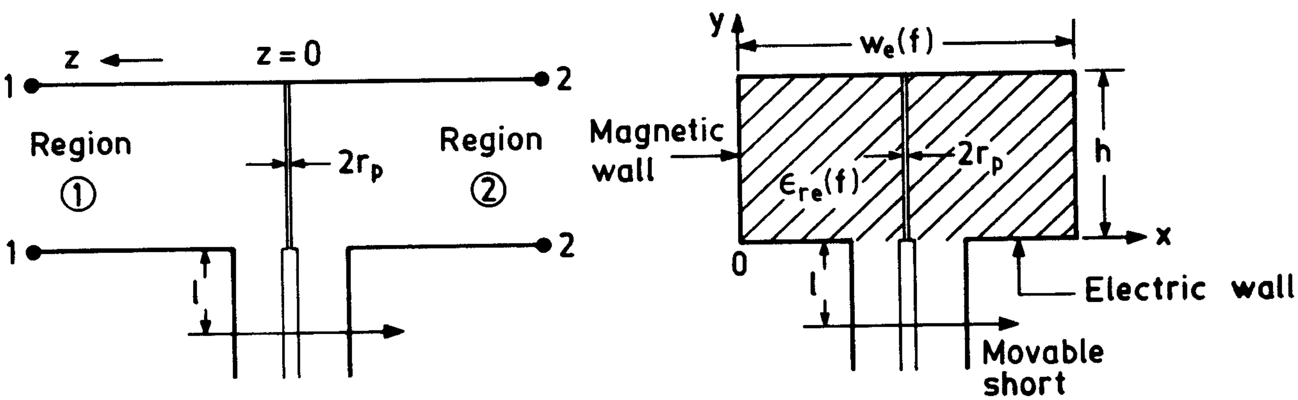

Fig. 4: Planar waveguide equivalence of the proposed structure.

A close look at the equivalent circuit shown in Fig. 5 shows that the reactance of the feeding probe has been modelled by an expression similar to the one used for writing the input reactance of a thin-wire antenna situated in an

*Such a behavior was first reported by the authors [2] in case of similar configurations using rectangular waveguide.

unbounded space. Also, the expression for Z_{op} (eq. (1)) tallies with the well-known expression for the characteristic impedance of a thin-wire antenna, of length h and of radius r_p, radiating into an homogeneous unbounded dielectric - field space. Thus we establish that, for the purpose of writing its input reactance, the feeding probe can be accurately modelled if we let it be equivelenced to a thin cylindrical antenna situated in an infinite homogeneous space with dielectric constant $\varepsilon_{re}(f)$.

Table I: Comparison between theoretical and experimental values of resonant length (r_p=0.41 mm, Z_{om}=50 ohms, $_r$=2.55, h=3.175mm, Z_o=55 ohms).

Frequency in GHz	Resonant length in mm: Theoretical	Resonant length in mm: Experimental
2.00	67.87	68.58
2.25	59.57	60.40
2.50	52.94	54.10
2.75	47.52	46.74
3.00	43.02	42.70
3.25	39.21	38.90
3.50	35.96	34.98
3.75	33.78	31.78
4.00	30.69	30.13

References

(1) H. Howe, Stripline Circuit Design, Dedham, MA: Artech House, 1974.

(2) R.S. Tomar and C. Das Gupta, 'Analysis and filtering applications of two newly-proposed waveguide-coaxial line junctions', Journal of Applied Phys., vol. 54, No. 8, pp. 4623-4628, August,1983.

(3) G. Kompa and R. Mehran, 'Planar Waveguide Model for Calculating Microstrip Components', Electronics Letters, vol. 11, No. 19, pp. 459-460, 18th September,1975.

Table II: Comparison between theoretical and experimental values of resonant length (r_p = 0.755 mm, Z_{om} = 40 ohms, $_r$ = 2.55, h=3.175 mm, Z_o = 55 ohms).

Frequency in GHz.	Resonant length in mm: Theoretical	Resonant length in mm: Experimental
2.00	71.95	69.82
2.25	63.61	62.39
2.50	56.94	55.84
2.75	51.48	48.40
3.00	46.92	43.81
3.25	43.07	39.74
3.50	39.76	37.72
3.75	36.89	34.04
4.00	34.33	30.60