TABLE VII
SELF-CAPACITANCES FOR UNCOUPLED MICROSTRIP DIELECTRIC CONSTANT OF 2.4

W/H	CG/2
4.0	.6147
3.50	.5596
3.0	.5040
2.5	.4477
2.0	.3904
1.6	.3436
1.2	.2954
1.0	.2705
.8	.2446
.6	.2173
.4	.1874
.2	.1514

microstrip filter does not have a spurious second harmonic response, and it takes up less space at the expense of shorts through the substrate. Microstrip filters are not viable by themselves because they have poor ultimate rejection, high loss, and do not follow the theoretical curves; but it is sometimes convenient and economical to use them.

V. Conclusion

In conclusion, a procedure is given which can be used to design coupled line and interdigital structures on microstrip. The procedure can be easily computer programmed using the polynomial approximations to give accurate results with very short computation times compared with times required by the Bryant and Weiss method or even Smith's approximations. Most importantly, since this procedure leads from electrical parameters to dimensions, it can be incorporated in automatic design programs.

References

[1] T. G. Bryant and J. A. Weiss, "Parameters of microstrip transmission lines and coupled pairs of microstrip lines," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-16, pp. 1021–1027, Dec. 1968.
[2] J. I. Smith, "The even and odd-mode capacitance parameters for coupled lines in suspended substrate," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-19, pp. 424–430, May 1971.
[3] R. J. Wenzel, "Exact theory of interdigital band-pass filters and related coupled structures," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-13, pp. 559–575, Sept. 1965.
[4] W. J. Getsinger, "Coupled rectangular bars between parallel plates," *IRE Trans. Microwave Theory Tech.*, vol. MTT-10, pp. 65–72, Jan. 1962.
[5] E. G. Cristal, "Coupled circular cylindrical rods between parallel ground planes," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-12, pp. 428–439, July 1964.
[6] G. L. Matthaei, "Interdigital band-pass filters," *IRE Trans. Microwave Theory Tech.*, vol. MTT-10, pp. 479–491, Nov. 1962.

Asymmetric Even-Mode Fringing Capacitance

HENRY J. RIBLET, FELLOW, IEEE

***Abstract*—An expression is given for the even-mode fringing capacitance of an infinite rectangular bar, asymmetrically located inside an infinite u-shaped outer conductor.**

Manuscript received May 3, 1976; revised August 30, 1976.
The author is with the Microwave Development Laboratories, Inc., Needham Heights, MA 02194.

Introduction

The writer [1] has recently given a closed expression for the odd-mode fringing capacitance for an infinite rectangular bar, asymmetrically located in an infinite u-shaped outer conductor. The essential problem solved in that note was the determination of the conformal transformation which maps the upper half t plane into the doubly infinite u-shaped polygon in the z plane as shown in Fig. 1. The determination of the capacitance of the structure presented no problem since it could be found from well-known formulas with the help of the "excess capacitance" introduced by Riblet [2].

If the even-mode capacitance is defined in a manner consistent with that used by Getsinger [3], as the capacitance of the structure in the z plane when the line segment BC is a magnetic wall, then we require in the t plane the capacitance of two separated line segments, AB and CD, both at the same potential, with respect to the infinite line segment DA. The determination of the limiting value of this capacitance is the essential problem of this short paper.

The Even-Mode Capacitance

In the t plane, the capacitance between the two-line segments, $[\mu + \delta\mu, 1]$ and $[1/k^2, \nu - \delta\nu]$ maintained at the same potential, and the infinite line segment $[\nu + \delta\nu, \mu - \delta\mu]$ is required in the limit as $\delta\mu$ and $\delta\nu \to 0$. It is important to keep in mind that the small semicircles about A and D are magnetic walls, while the semicircles about O, B, C, and E play no essential role in the calculations. This capacitance is not altered if the upper half of the t plane is mapped onto the upper half of the s plane so that B maps into -1, C into $+1$, A into $-l$, and D into $+l$. This is accomplished by the linear transformation

$$s = \gamma \frac{t - \alpha}{t - \beta} \tag{1}$$

if α, β, and γ are selected so that

$$\gamma \frac{1/k^2 - \alpha}{1/k^2 - \beta} = -\gamma \frac{1 - \alpha}{1 - \beta} = 1 \tag{2}$$

and

$$\gamma \frac{\nu - \alpha}{\nu - \beta} = -\gamma \frac{\mu - \alpha}{\mu - \beta} = l. \tag{3}$$

Again it is important that the semicircles about $-l$ and $+l$ and the line segment between B and C be magnetic walls. From (2) and (3)

$$\begin{aligned} 2\alpha\beta - (1 + 1/k^2)(\alpha + \beta) + 2/k^2 &= 0 \\ 2\alpha\beta - (\mu + \nu)(\alpha + \beta) + 2\mu\nu &= 0. \end{aligned} \tag{4}$$

Whenever $\mu + \gamma \neq 1 + 1/k^2$, this set of equations can be solved uniquely for $\alpha\beta$ and $\alpha + \beta$. It is then a simple matter to solve the quadratic equation

$$\chi^2 - (\alpha + \beta)\chi + \alpha\beta = 0$$

to determine α and β. Gamma is then found from either (2) or (3). The total capacitance of the system is unchanged by the transformation, and, if we take the radii of the semicircles about A and D in the s plane to be the same, the geometry and the lines of force in the s plane are completely symmetrical about the imaginary axis. Thus it may be replaced by a magnetic wall. Then one-half of the limiting value of the total capacitance of the system is given by the limiting value of the capacitance of the finite line segment, $[1, l - \delta s]$, with respect to the infinite

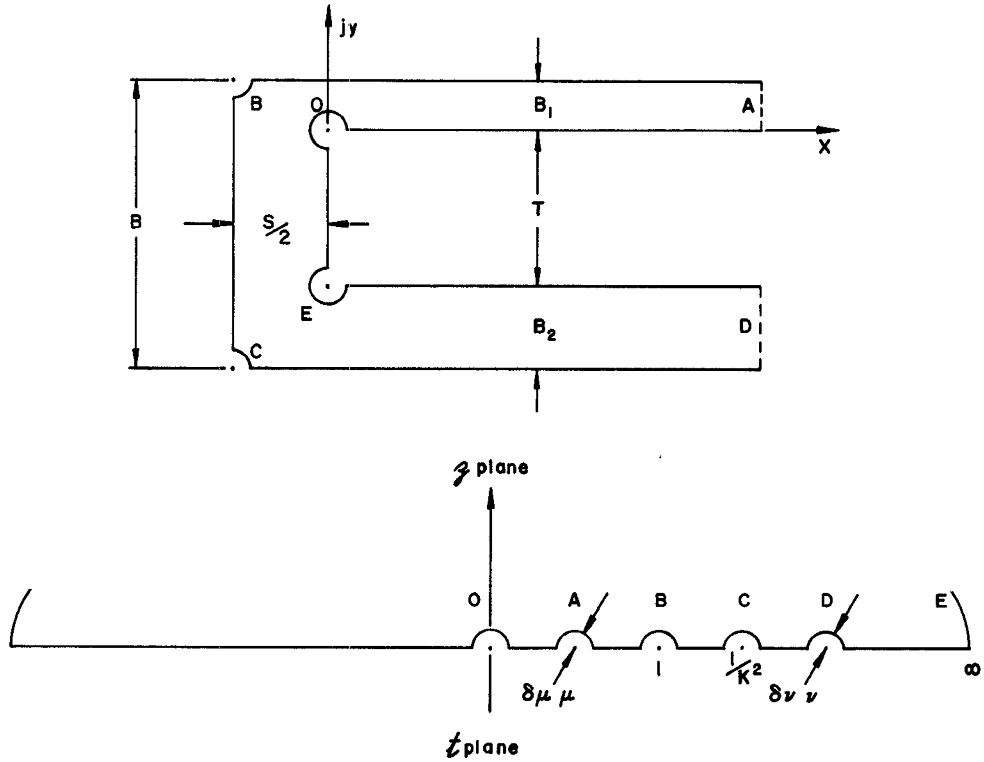

Fig. 1. The t and z planes.

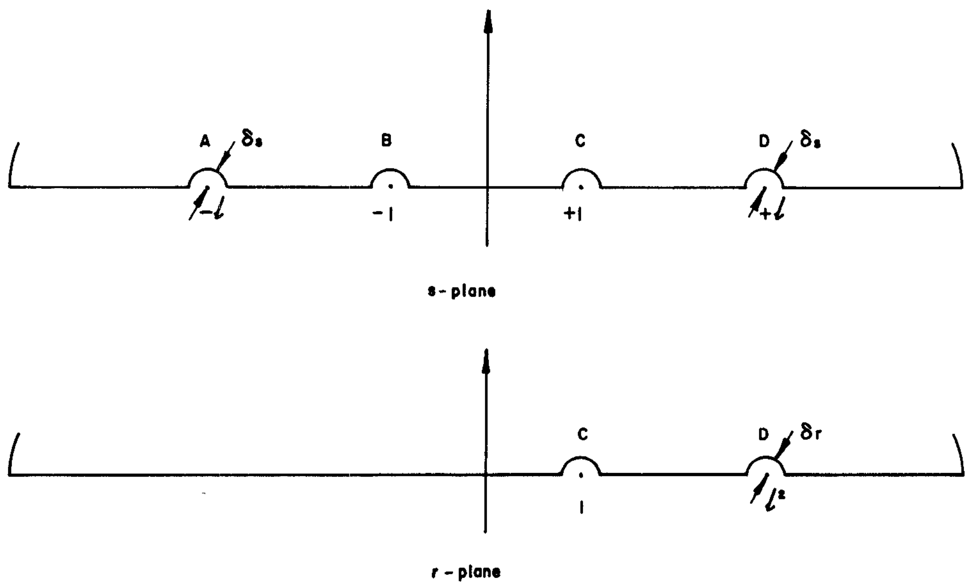

Fig. 2. The r and s planes.

line segment, $[l + \delta s, \infty]$, as $\delta s \to 0$, keeping in mind that the positive imaginary axis, the line segment, $[0,1]$, and the small semicircle about D are magnetic walls. The value of this capacitance can be readily found by mapping the upper right-hand quadrant of the s plane onto the upper half of the r plane (Fig. 2) by means of the transformation, $r = s^2$. Here the points on the negative real axis correspond to the points of the positive real axis of the s plane. The problem is thus reduced to finding the capacitance of the line segment, $[1, l^2 - \delta r]$, with respect to the infinite line segment, $[l^2 + \delta r, +\infty]$, in the r plane in the limit as $\delta r \to 0$. Here the line segment, $[-\infty, 1]$, and the small semicircle about D are magnetic walls.

Riblet [3] has shown how this capacitance, C_e', differs only by an "excess capacity," $\log (2)/\Pi$, from the capacitance obtained by mapping the upper half of the r plane into the interior of a rectangle. This latter capacitance is given by K'/K for $K^2 = (b - a)(d - c)/(d - b)(c - a)$, where $a = \to\infty$, $b = 1$, $c = l^2 - \delta r$, and $d = l^2 + \delta r$. Then $k^2 = 2\delta r(l^2 - 1)$ if higher powers of δr are neglected. Finally,

$$\frac{K'}{K} \approx \frac{1}{\Pi} \log \frac{16}{k^2} = \frac{1}{\Pi} \log 8 \frac{(l^2 - 1)}{\delta r} \tag{5}$$

so that

$$C_e' = \frac{K'}{K} - \frac{\log (2)}{\Pi} = \frac{1}{\Pi} \{\log 4(l^2 - 1) - \log \delta r\}. \tag{6}$$

Hence the total capacitance C_e of the system in the s plane in terms of the radii of the small semicircles in the s plane is

$$C_e = \frac{2}{\Pi} \{\log [2(l^2 - 1)/l] - \log \delta s\} \quad (7)$$

since $\delta r = 2s\delta s$.

By definition, the even-mode fringing capacitance, C_{f_e}'', is given by the limiting value of $C_e - C_{P_A} - C_{P_D}$, where C_{P_A} and C_{P_D} are the parallel-plate capacitances associated with the gaps B_1 and B_2 of Fig. 1. In [1], C_{f_o}'' was defined as $C_o - C_{P_A} - C_{P_D}$ and, since the formulas for the parallel-plate capacitances are rather involved, it is convenient to express C_{f_e}'' in terms of C_{f_o}''. Thus

$$C_{f_e}'' = C_{f_o}'' + C_e - C_o. \quad (8)$$

From [1, eq. (12)]

$$C_o = \frac{1}{\Pi} \{2 \log (\nu - \mu) - \log \delta\mu - \log \delta\nu\}. \quad (9)$$

Now $\delta\mu$ and $\delta\nu$ are obtained from δs with the help of (1). After differentiating, it is found that

$$\delta\mu = \frac{(\mu - \beta)^2}{\gamma(\alpha - \beta)} \delta s \quad (10)$$

and

$$\delta\nu = \frac{(\nu - \beta)^2}{\gamma(\alpha - \beta)} \delta s. \quad (11)$$

Then from (7) and (9)

$$C_o - C_e = \frac{2}{\Pi} \{\log (1 - k^2 \operatorname{sn}^2 a \operatorname{sn}^2 d) - 2 \log (k \operatorname{sn} d) - \log \frac{2(l^2 - 1)}{l} - \log \frac{(\mu - \beta)(\nu - \beta)}{\gamma(\alpha - \beta)}\}. \quad (12)$$

This expression, together with [1, eq. (13)], in view of (8) gives the desired formula for C_{f_e}''.

REFERENCES

[1] H. J. Riblet, "Asymmetric odd-mode fringing capacitances," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-24, pp. 158–159, Mar. 1976.
[2] ——, "The determination of an excess capacitance," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-22, p. 467, Apr. 1974.
[3] W. J. Getsinger, "Coupled rectangular bars between parallel plates," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-10, pp. 65–72, Jan. 1962.

An Analytical Comparison of Two Simple High-Q Gunn Oscillators

IAN D. HIGGINS AND ROBERT DAVIES

Abstract—**This note compares and analyzes two commonly used simple waveguide Gunn oscillators in terms of their loaded Q-factors. Suitable design criteria are established for both, and two oscillators which were tested conformed well to these. It is concluded that although the more mechanically complex oscillator, which is in common use, has a greater flexibility, the simpler oscillator is adequate for most applications.**

I. INTRODUCTION

The Gunn diode is a simple two-terminal device which, when mounted in a resonant circuit and biased with a suitable dc potential, generates microwave power. The basic noise and stability properties of the device are modified by the loaded Q-factor (Q_L) of the resonant circuit and for many applications a desirable value of Q_L is between 200 and 1000. Resonant circuits, or cavities, for this purpose are usually made from simple waveguide and Figs. 1 and 2 show two common types. The purpose of this short paper is to analyze the critical design aspects of these cavities and to determine if either has any basic advantages.

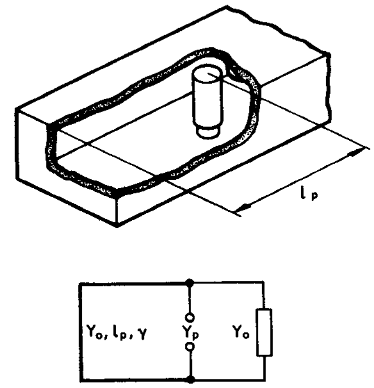

Fig. 1. Post-type oscillator and its equivalent circuit.

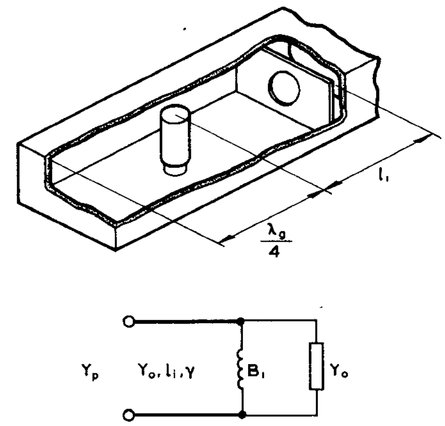

Fig. 2. Iris-coupled oscillator.

The oscillator shown in Fig. 1 has been previously studied [1] and there are many commercial samples of this type. It consists simply of a post-mounted Gunn diode spaced a half wavelength from a short circuit.

The second oscillator, which is mechanically more complex, consists of a Gunn post assembly mounted between a simple inductive (i.e., circular hole) iris and a waveguide short circuit, Fig. 2. There are also many oscillators of this design commercially available and it is commonly supposed [2] to have advantages over the more simple post-coupled oscillator.

Although the two oscillators appear simple in construction, the analyses are complex. A numerical analysis of the post-mounting structure was given by Eisenhart and Kahn [3] in 1971, and this analysis is used for final evaluation of both oscillators. However, a more basic analytical approach is adopted in this short paper in order to give a meaningful comparison of the two cavities. The simplified analysis is only concerned with the circuit external to the post since complex effects of the post are the same for both circuits. In this respect it differs from the analysis of White [4] and leads to simple expressions for the oscillators' Q-factors. The interface reference plane is at the waveguide/post junction Y_p representing the admittance seen by the "Gunn–package–post." The equivalent

Manuscript received March 12, 1976; revised August 23, 1976.
The authors are with the Mullard Research Laboratories, Cross Oak Lane, Redhill, Surrey RH1 5HA, England.