TABLE II

FACTOR G BY TWO METHODS

				G	
Line	Mode	ϵ_{ei}	f_i	Inflection Point	Equation (3)
1b	even	8.0	11.4	0.87	0.87
2b	even	8.0	12.8	0.835	0.89

lines, also hold for the even and odd modes of parallel-coupled microstrip lines if total mode impedances are employed in the formulas.

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Effects of the T-Junction Discontinuity on the Design of Microstrip Directional Couplers

R. W. VOGEL

Abstract—**The influence of the T-junction effect on the microstrip 3-dB branchline and rat-race directional couplers design has been described. In the proposed method, the T-junction equivalent circuit similar to the circuit used by Franco and Oliner [1] for the symmetric stripline junction is applied. Simple design expressions that combine the modified characteristic impedance and electrical length of the coupler arm with the equivalent parameters of the T junction are given. These expressions are helpful in the determination of the microstrip coupler dimensions. The experimental verification of an improvement of the design procedure is also included.**

In a paper by Leighton and Milnes [2], the influence of T-junction reactances on the characteristics of 3-dB branch line and rat-race couplers has been considered. The method proposed in [2] was based on the T-junction equivalent circuit used by Altschuler and Oliner [3]. In the practical design of all microwave devices containing the T junctions, it is sometimes more convenient to represent this discontinuity by the equivalent circuit in the form recommended by Franco and Oliner [1] (Fig. 1). Knowing the parameters of this equivalent circuit, we can determine the correct way of finding the main dimensions of the devices that contain the T junction.

In order to design the 3-dB branch line and rat-race directional couplers, the calculations based on the procedure recommended by Franco and Oliner were carried out, with the difference being that the effective linewidth D was determined with the same procedure as by Leighton and Milnes [2]. The results obtained in the range of impedance ratios usually met (in the case of 3-dB branch line and rat-race couplers) are presented in a graphical form (Figs. 2 and 3) showing the dependence of the circuit parameters on the geometrical dimensions of the lines involved and on the frequencies used. As mentioned earlier, the parameters can be applied to the design of some branch and ring-type 3-dB couplers. This design is possible when one compares two equivalent circuits, of which the first represents the ideal coupler without junction reactances and the second corresponds to the coupler, including the parameters of the T junction. These equivalent circuits can be analyzed by the even-and-odd excitation method employed by Reed and Wheeler [4]. Determining and comparing the wave transmission matrix for both cases, we can evaluate the modified characteristic impedance Z_m of the branch (stub arm) and its electrical length Θ_m between the reference planes. These new parameters compensate discontinuity effects of the T junction. For

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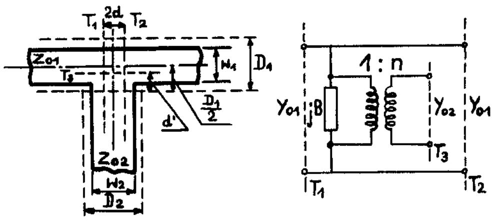


Fig. 1. Equivalent circuit for microstrip T junction.

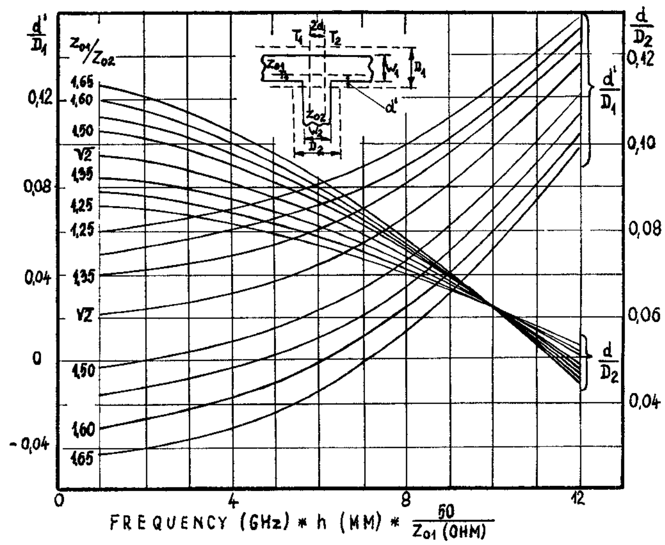


Fig. 2. Reference plane shifts in the main and stub arms as a function of frequency f and substrate thickness h.

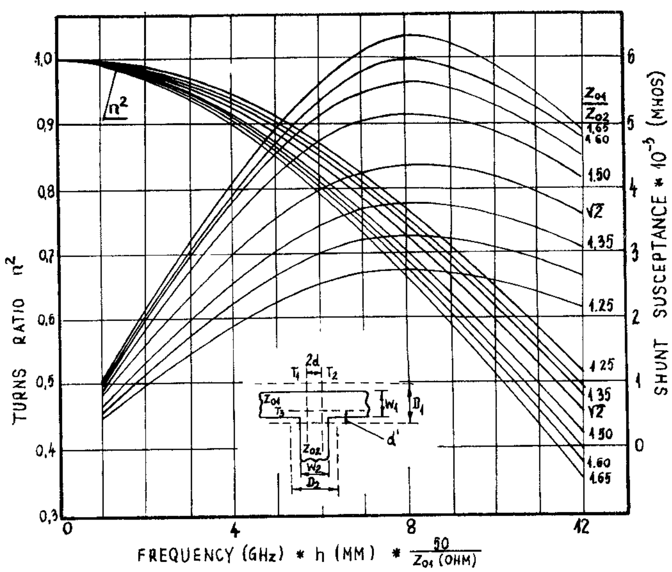


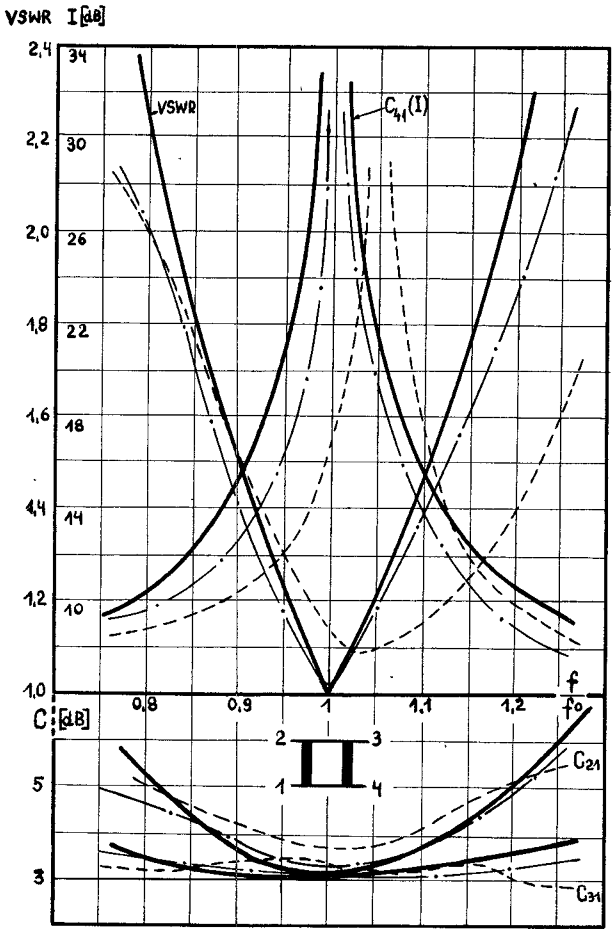
Fig. 3. Transformer turn ratio n and shunt susceptance B as a function of frequency f and substrate thickness h.

the case of the branch line coupler, they are given by

$$Z_m = \frac{n^2 Z_i}{\sqrt{1 - (BZ_i)^2}}$$
$$\Theta_m = \arccos (BZ_i) \tag{1}$$

where

Z_i characteristic impedance of branch (stub arm) for the ideal case without junction effects (for the 3-dB coupler, $Z_i = Z_0/\sqrt{2}$);



(a)

(b)

Fig. 4. Comparison between idealized theory and experimental results for a two-branch 3-dB coupler. ———: idealized theory, without junction effects; ---------: experimental results for the idealized design; ---·---·---: experimental results for the improved design method.

n, B parameters of the equivalent circuit of the T junction, the transformer turns ratio and shunt susceptance, respectively.

For the rat-race coupler, the modified characteristic impedance of the ring and the electrical length of the $\lambda/4$ and $\frac{3}{4}\lambda$ sections can be expressed as

$$Z_m = \frac{Z_i}{n^2 \sqrt{1 - \left(\frac{BZ_i}{2n^2}\right)^2}}$$

$$\Theta_{m1} = \arccos\left(\frac{BZ_i}{2n^2}\right), \qquad \text{for the } \lambda/4 \text{ section}$$

$$\Theta_{m2} = \pi + \arccos\left(\frac{BZ_i}{2n^2}\right), \qquad \text{for the } \tfrac{3}{4}\lambda \text{ section} \tag{2}$$

where

Z_i characteristic impedance of the ring for the ideal case (for the 3-dB coupler, $Z_i = \sqrt{2}Z_0$);

n, B parameters of the equivalent circuit of the T junction.

The results obtained from the (1) and (2) are not the final parameters of the couplers because, primarily, only Z_i is known and the first computation gives the impedance of a new T junction, which must be considered in the next step. The final parameters Z_{mf} and Θ_f can be found by successive approximation, and in practice two or three iterations are usually sufficient. Knowing the final parameters, the proper location of the reference planes T_1, T_2, T_3 given by the d and d' can be determined.

In order to test the usefulness of the theory, S-band two-branch and rat-race 3-dB couplers were constructed in microstrip. In Fig. 4(a) and (b), the VSWR, isolation, and coupling of two-branch couplers are plotted. Similar experiments for the rat-race couplers were made, but in this case the junction effects were much less significant than in the two-branch couplers. In all cases, the couplers were made on a substrate of 0.5-mm thick alumina, having a dielectric constant of 9.7. The nominal center frequency was 3 GHz.

It is worth noting that experimental directional couplers that were designed according to the procedure developed have performance characteristics in much better agreement with theoretical ones than devices designed without considering the T-junction effect.

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Precise Frequency and Phase Control of LSA Oscillators

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***Abstract*—It is shown that consideration of both the external microwave circuit and the dc bias applied to the diode are necessary in order to obtain satisfactory locking results with an LSA oscillator. When both are properly arranged, good control of both the operating frequency and the phase of an LSA oscillator at 3 GHz is demonstrated.**

Well-behaved frequency and phase-locking characteristics for LSA oscillators can be achieved only after careful attention has been paid to the particular operating parameters of these devices. As has been pointed out by Jeppesen and Jeppson [1] and Camp [2], the LSA mode of oscillation is fundamentally a voltage-controlled relaxation mode. As such, the frequency and spectral purity of the output from these devices is strongly dependent upon the characteristics of the applied bias voltage. Secondly, because the high-power LSA device has an active impedance on the order of 100 Ω, high dc-to-RF conversion efficiency can be achieved only through very heavy coupling between the device and the commonly employed 50-Ω load circuit. This heavy coupling makes the performance of the oscillator very susceptible to external circuit variations. Therefore, careful control of the bias pulse applied to the oscillator, as well as the load characteristics of the external circuit, are necessary before good locking performance can be observed.

The microwave circuit in which the present experiments were carried out is shown in Fig. 1. It is a multiaxis radial circuit (MARC) first proposed by Eastman [3].

Camp [2] has shown that for optimum LSA efficiency, the second harmonic should be reactively terminated with a high-impedance inductive load. Because the length of the shorted radial line in the MARC is slightly less than an eighth wavelength at the fundamental frequency, the second harmonic load created by the cavity is automatically correct. To insure that the output line is also properly tuned for the second harmonic, a method also reported by Eastman [4] is employed. A low-impedance tuning slug, an eighth of a wavelength long at the fundamental frequency, is inserted in the output coaxial line. It is located approximately one-quarter wavelength from

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