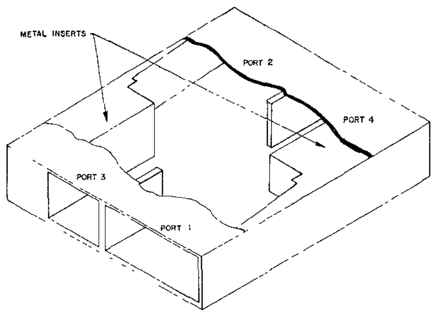

Fig. 2. Zero-dB coupler.

0.15 dB. No loads were required on ports 3 and 4, and no external matching was necessary.

A third application of the principle would consist of a compact round-to-rectangular waveguide transition.

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Pulse Power Capacity of Short-Slot Couplers

A short-slot sidewall coupler and a short-slot topwall coupler were both tested at S band for their pulse power capacity. Both types of coupler were tested in the 0-dB coupler configuration [1], which had the practical convenience of requiring only one high-power load.

In order to make meaningful tests, a special test section of S-band waveguide was constructed. This test section had the standard WR-284 waveguide width of 2.840 inches, but was only 0.447-inch high, which is one third of the standard waveguide height of 1.340 inches. A two-section waveguide transformer was placed at each end of the test section to match it to standard waveguide; the test section, including both transformers, was measured to have a maximum VSWR of 1.03 over the frequency band 2.7 to 2.9 Gc/s. All the measurements were made at a frequency of 2.856 Gc/s. The VSWR of the high-power water load at this frequency was better than 1.05. The pulse length used throughout the tests was 3 μs.

The test procedure in each case was to increase the klystron power until arcing occurred in the waveguide. The klystron power was then reduced until arcing stopped, and this power level was maintained for 10 minutes without any further arcing. Taking the test section, the sidewall coupler, and the topwall coupler in turn, the all-clear pluse power was 1.25, 2.72, and 1.67 MW, respectively. (The corresponding all-clear power averages were 0.431, 0.94, and 0.576 kW, respectively.)

Given the fact that the test section is only one third of the standard waveguide height, the pulse power capacity of the short-slot sidewall coupler is approximately 72 per cent of WR-284, and the pulse power capacity of the short-slot topwall coupler is approximately 44 per cent of WR-284 waveguide.

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Accurate Measurements for Dynamic Representations of Parametric Amplifier Varactor Diodes

Varactor diode parameter representations useful for describing circuit performance are usually derived from UHF or microwave measurements. Although many measurement methods have been used to date, the two main classes appear to be measurements which result in "static" and "dynamic" diode representations.

"Static" representations are defined here as those in which a UHF or microwave signal of small amplitude is used to measure characteristics of the diode which can then be expressed in terms of an equivalent circuit. UHF bridge and Q-meter [1], [2], microwave junction [3], and microwave cavity [4] measurements have been used to obtain equivalent circuits. Smith-chart data obtained from measured microwave VSWR data obtained as a function of diode bias [5], [6] or frequency [7] has been used to arrive at equivalent circuit parameters for the diode junction or cartridge. Measurements yielding circuit parameters as polynomial functions of frequency have been developed [8].

"Dynamic" representations are defined as those which characterize the diode in terms of a parameter which expresses the diode's performance under pumping conditions. It is the purpose of this note to propose a rapid and accurate measurement scheme for obtaining a dynamic quality parameter applicable to the commonly used parametric amplifier varactor.

A dynamic parameter ω_d is defined as the ratio of the total elastance variation to four times the series resistance of a pumped parametric amplifier diode [8]: $\omega_d = \Delta(1/C)/4R$. The extremes of elastance variation correspond to definite values of diode capacitance, such that if C_1 and C_2 are respectively the minimum and maximum diode capacitance, ω_d becomes

$$\omega_d = 2\pi f_d = \frac{1}{4R}\left(\frac{1}{C_1} - \frac{1}{C_2}\right) \tag{1}$$

where f_d is the "quality parameter." Let diode impedances Z_1 and Z_2 correspond to diode conditions which produced capacitances C_1 and C_2. If Z_1 and Z_2 are valid at frequency f, then

$$1 + j\,(f/2f_d) = (Z_1 + Z_2^*)/(Z_1 - Z_2) \tag{2}$$

where the asterisk denotes the complex conjugate. The measurement of Z_1 and Z_2 at the diode is not so easily accomplished, but measurements may be made when the diode terminates a transmission line. Assume that a lossless transformer is inserted between the diode and the transmission line. At the input to the transformer, the reflection coefficients Γ_1 and Γ_2 corresponding to impedances Z_1 and Z_2 may then be defined at frequency f, and f_d may be written [9] as

$$f_d = \frac{f}{2}\left[\left|\frac{1 - \Gamma_1\Gamma_2^*}{\Gamma_1 - \Gamma_2}\right|^2 - 1\right]^{-1/2} \tag{3}$$

Equation (3) can be rewritten in terms of the magnitudes γ_1 and γ_2, and of the phase difference Ψ between Γ_1 and Γ_2:

$$f_d = \frac{f}{2}\left[\frac{\gamma_1^2 - \gamma_2^2 - 2\gamma_1\gamma_2\cos\Psi}{(1-\gamma_1^2)(1-\gamma_2^2)}\right]^{1/2} \tag{4}$$

Now if the lossless transformer is adjusted such that either γ_1 and γ_2 is zero, (4) can be simplified and made independent of the phase Ψ; setting $\gamma_2 = 0$ results in

$$f_d = \frac{f}{2}\left[\frac{1}{\gamma_1^2} - 1\right]^{-1/2} \tag{5}$$

The measurement of f_d is now reduced to the measurement of a single reflection coefficient magnitude at a single frequency. For practical values of f, f_d is usually large ($f_d \gg f$) such that γ_1 is large and difficult to measure accurately using standard slotted-line techniques. However, if a reflectometer is used

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