The probability of two waveguides being aligned with only a or only b displacements is therefore quite small. An additional consideration is that one is usually concerned with the overall return loss of several flanges in a waveguide run, and the probability that all flanges will be misaligned similarly is even more remote.

The original criterion of basing the initial aligned return loss on the case where the waveguides have extreme tolerances $a + \Delta a, b - \Delta b$ and $a - \Delta a, b + \Delta b$ is also a statistically remote condition. Usually, the return loss will be much better. In fact, in the case where the two waveguides have a negligible impedance discontinuity, the return losses for the displacements shown in Table II are 6 dB better. Alternatively, one can state that in this case the deviation required to give the return loss values allowed in Table II is approximately $0.025a$, a figure significantly larger than that of (18), i.e.,

$$\Delta a' = 0.025a \text{ (matched-waveguide case).} \tag{19}$$

In practice it is found that while it may be quite difficult to dimension the flanges to ensure that (18) is not exceeded, an allowable displacement approximately midway between the $0.0175a$ of (18) and the $0.025a$ of (19) is more readily feasible. Such a compromise value of $0.21a$ gives a worst return loss of approximately -40 dB. It makes reasonable allowance for the statistical considerations, enables reasonable tolerances to be assigned in most instances, and is essentially in accordance with the basic logic behind the IEC 1-dB degradation criterion.

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A 60-W CW Solid-State Oscillator at C Band

R. N. WALLACE, M. G. ADLERSTEIN, AND S. R. STEELE

***Abstract*—A 60-W CW solid-state oscillator has been developed for operation in C band. The oscillator combines the power of six high-efficiency GaAs multimesa Read diodes. Single-diode oscillators have given power outputs as high as 13.3-W CW at 5 GHz.**

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A solid-state oscillator delivering 60-W nominal CW power in C band has been constructed and tested. The oscillator circuit combines the outputs from six high-efficiency multimesa GaAs Read IMPATT diodes. Such diodes, incorporating chip-level power combining, allow one to reach the 60-W level with relatively few discrete devices. Compared to an oscillator combining a large number of low-power diodes, the source described here operates with simpler bias circuitry and is easier to tune for optimum performance.

Fabrication of the high-power diodes from low-high-low Read profile epitaxial GaAs wafers grown in our laboratory has been described in some detail elsewhere [1]. Individual diodes consist of four separate mesas in a 2×2 array, mounted on an integral plated-gold heat sink. This arrangement provides a substantial improvement in thermal resistance over that obtained with a single mesa of equivalent area [2], and we regularly measure values of 4.5–5.0°C/W in C-band diodes. With a room-temperature heat sink, a diode having 22–23-percent efficiency can thus produce 10-W CW output with a junction temperature of ~ 200°C. The plated heat-sink technology is suited for large-volume, low-cost diode production, and offers economic advantages over the IIa diamond heat sinks often used in obtaining high-power operation [3].

The yield of diodes producing 10-W CW or more is not yet large, but our results indicate that the devices will be manufacturable. In a recent series of thirteen epitaxial wafers selected for processing, eleven produced diodes which, when operated with 40°C (nominal) heat-sink temperature, reached or exceeded 10-W CW output. Efficiencies were typically 22–25 percent, and frequencies of optimum performance ranged from 4.7 to 6.5 GHz. Two wafers, grown in different epitaxial reactors, produced best results of 13.3-W CW output with 24-percent efficiency in the 4.8–4.9-GHz range. A 15.3-W CW result with a cooled heat sink was reported previously [1]. These are among the highest power outputs reported for C-band IMPATT diodes.

The six-diode oscillator circuit used in the present work is similar to the type described by Harp and Stover [4]. The circuit concept has been analyzed in considerable detail and tested experimentally by Kurokawa [5], [6]. A few practical operating considerations will be noted here.

Fig. 1 is a schematic representation of the oscillator circuit. The basic resonator is a cylindrical TM_{010}-mode cavity. The cavity frequency is adjusted with a dielectric rod tuner, and coupling to the external load is controlled by varying the penetration of the coaxial output probe. The six diodes are coupled to the main resonator through coaxial lines ($Z_0 = 50\ \Omega$) passing along the cavity side wall. Coupling between the diodes and the cavity is adjusted by moving the diode mounting plugs axially, and by changing the dimensions of the individual slug transformers. Bias is supplied to the diodes along the coaxial center conductors, which pass through absorbing terminations at the top of the cavity. The circuit is water cooled during operation.

The operation and tuning of the circuit can be conveniently described in terms of the impedance Z_m measured on the diode coaxial lines at the midplane of the cavity. For frequencies near resonance, this impedance is essentially that of a parallel RLC circuit (the loaded cavity) in series with a resistor (the terminated bias line). At resonance, Z_m is resistive and may range from $\sim 1.5Z_0$ to $\sim 20Z_0$ depending on the output coupling adjustment.

Large-signal terminal impedances of the individual diodes are approximately $-0.8 + j6\ \Omega$ near 5 GHz. The slug and coaxial line must thus be designed to transform Z_m to the much lower impedance required by the diode. The range of adjustments

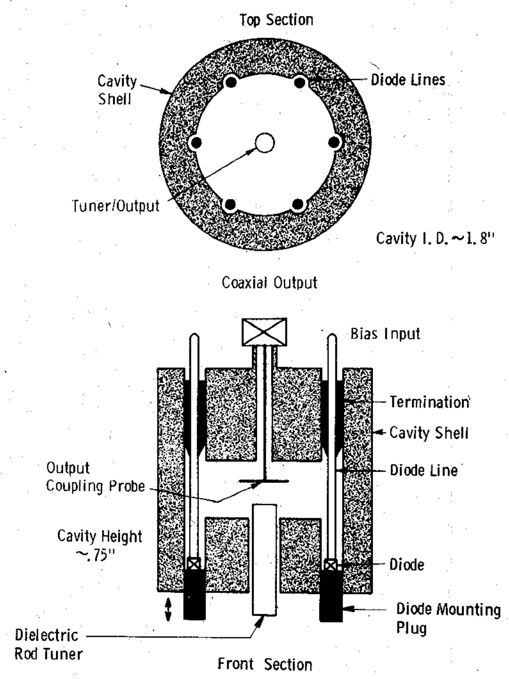

Fig. 1. Schematic drawing of the cavity-type power combiner.

possible in the circuit allows one to match the diode with many different combinations of Z_m and slug dimensions. However, a choice of Z_m giving good efficiency must be made. Small values of Z_m result in excessive losses to the bias line termination.[1] Large values of Z_m cause much of the available diode power to be dissipated in cavity-wall losses. For the present circuit, intermediate values of Z_m, typically $6Z_0$–$7Z_0$ at resonance, were found to be most satisfactory.

Final adjustment of the oscillator is greatly aided by the fact that each diode serves as its own tuning indicator. When operated with a high-impedance bias source, as is required for stabilization against bias oscillations [7], the diode voltage for a given bias current is a strong function of the RF load impedance. The particular voltage V_0 corresponding to optimum tuning is determined for each diode in single-diode tests. In the six-diode circuit, an operating voltage above V_0 for a particular diode indicates that the real part of its RF load impedance is too large. Voltages below V_0 indicate load impedances (real part) smaller than optimum. The slug transformers can be adjusted appropriately to correct these conditions.

Table I summarizes the performance of the six-diode oscillator at various bias current levels. Power output was peaked by adjusting the tuner and output coupling for each test condition. The diodes used averaged 11.2-W output with 22.6-percent efficiency in single-diode tests. The 62-W CW result with 20.9-percent efficiency at 5.12 GHz thus implies a combining efficiency of ~93 percent. Overall efficiency was fairly constant with changes in bias, as was the case with single diodes [1]. This means that power level can be reduced to extend lifetime without a severe efficiency penalty.

The oscillator could be mechanically tuned over a band ~100 MHz wide with less than 1-dB power degradation by adjusting only the dielectric rod tuner. Maintaining the power over wider bandwidths requires readjustment of both the output

[1] By using terminations which are intentionally mismatched (VSWR ~ 2), RF losses to the bias lines can be reduced. Excessive mismatch will lead to undesired instabilities, however.

TABLE I
MICROWAVE PERFORMANCE OF THE SIX-DIODE POWER COMBINER

P_{DC} (W)	P_{RF}* (W)	f (GHz)	η (percent)	Nominal Current per Diode (mA)
297.2	62.0	5.120	20.9	500
271.0	54.7	5.121	20.2	450
240.5	48.0	5.122	20.0	400
209.4	40.8	5.120	19.5	350
180.1	33.6	5.120	18.7	300
151.4	26.3	5.130	17.4	250

* Returned for maximum output at each dc input power level.

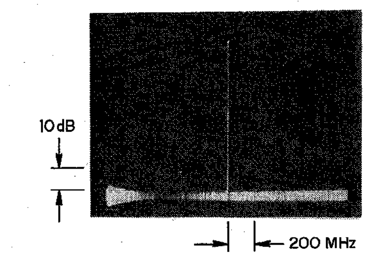

Fig. 2. Output spectrum of the six-diode oscillator in the free-running mode. f = 5.089 GHz, P = 60-W CW (nominal). Horizontal: 200 MHz/div. Vertical: 10 dB/div.

coupling and the diode positions. Electronic tuning of the source was also demonstrated by operating it as an injection-locked oscillator. The locking behavior was conventional, but the bandwidth was narrow, about 6 MHz with 300-mW injected signal, because of the relatively large external Q of the circuit.

The RF spectrum of the oscillator (Fig. 2) was a clean single line, free from spurious outputs. Frequency stability was relatively good even without injection locking. After warmup, the frequency remained within 1 MHz of its initial value for extended periods of laboratory operation. Detailed measurements of AM and FM noise remain to be made. However, noise in the multiple-diode oscillator is expected to be of the same order as that measured for single high-power high-efficiency GaAs Read diodes [5]. Such diodes typically have FM noise measures of 45–50 dB, and AM carrier-to-noise ratios exceeding 140 dB (1-Hz bandwidth) for frequencies more than 1 kHz from the carrier.

Turn-on and turn-off of the oscillator, once tuning was completed, was accomplished by slowly applying or removing the bias current on all diodes simultaneously. No retuning was required. Oscillation started when the bias current reached ~50 mA/diode and continued as a single line without mode hopping as current was increased to the normal operating level, ~500 mA/diode. Total frequency excursion during turn-on or turn-off was less than 10 MHz, and no bias oscillations were observed. This tractable behavior over a 10 to 1 range in bias currents leads us to expect only smooth variations in power output and frequency as the temperature of the oscillator is varied.

In conclusion, a 60-W CW 5-GHz oscillator combining six GaAs Read IMPATT diodes has been demonstrated. The behavior of the oscillator is tractable, and the use of a relatively small number of high-power diodes simplifies tuning and operation. The diodes employ a multimesa plated-heat-sink geometry suitable for mass production.

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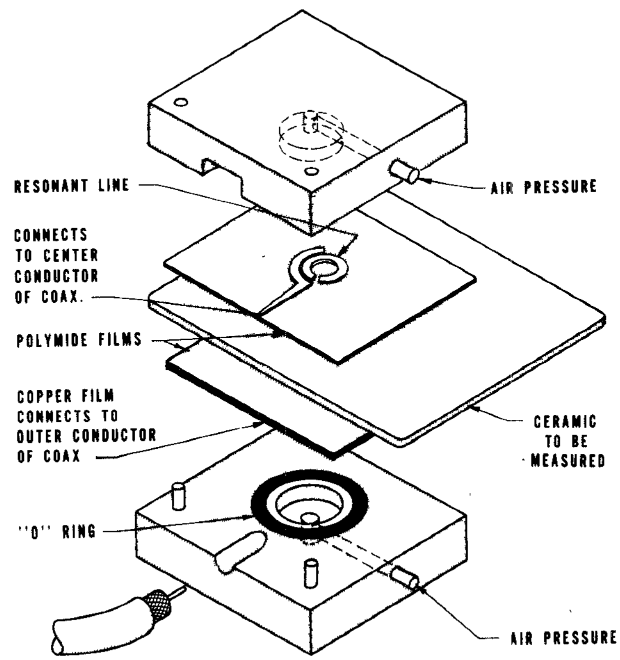

Fig. 1. Fixture for measuring dielectric constant of substrates for microstrip applications.

Measuring Dielectric Constant of Substrates for Microstrip Applications

A. R. GERHARD

Abstract—**A new technique for measuring the dielectric constant of unmetallized ceramic substrates for microstrip applications is fast, accurate, and nondestructive. Measurement is made at the actual microwave frequency at which the ceramic will be used. Results are repeatable to within ±0.1 percent of the dielectric constant relative to a known standard substrate. A measurement rate of 100/h can easily be achieved. A circuit is described which is used at 1.4 GHz and measures an area of approximately 1/2-in diameter on 25-mil-thick alumina substrates.**

Introduction

Variations in the dielectric constant of ceramic substrates intended for use in microwave integrated circuits often fall outside the stringent limits required to meet circuit performance specifications. Typical specifications call for alumina substrates 24–26 mils thick with a dielectric constant of 9.9–10.1. Therefore, in the interests of economy, it is desirable to measure the dielectric constant of the ceramic before the generation of a circuit and the application of devices. Available methods for achieving such meausrements such as the "two-fluid method" [1] and the "resonant-substrate method" [2]–[6] are not production oriented either because of the long time required to make a measurement or the fact that the method is destructive, preventing use of the material after the measurement.

Technique

The technique described here involves temporarily holding a conductor pattern against one side of the ceramic being measured and a ground plane against the opposite side of the ceramic. A swept microwave signal, with a mean frequency approximately the same as the frequency at which the final circuit will be used, is coupled to the resonant line. Reflected signals are observed, and the resonant frequency (as indicated by a sharp dip in reflected power) is measured. By comparison with a standard substrate, the dielectric constant of the unknown may be easily obtained by using the following relationships:[1]

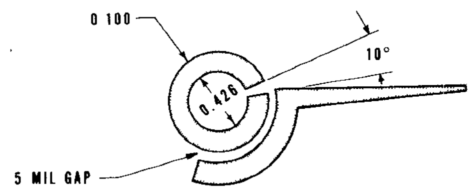

Fig. 2. Resonant circuit and coupling line.

$$\frac{\Delta\varepsilon_r}{\varepsilon_r} = K\frac{\Delta f_{\text{res}}}{f_{\text{res}}} \tag{1}$$

where

$\dfrac{\Delta\varepsilon_r}{\varepsilon_r}$ percent of change in dielectric constant;

$\dfrac{\Delta f_{\text{res}}}{f_{\text{res}}}$ percent of change in resonant frequency;

K −2.15 for the fixture shown in Fig. 1 when measuring alumina substrates having a nominal thickness of 25 mils [7]. The K factor is a function of linewidth, substrate thickness, frequency, and the dielectric materials used in the fixture [8], [9].

Fig. 1 shows typical fixturing that may be used with the resonant-line technique. The ceramic to be measured is placed between two films of 1-mil polyimide (such as DuPont KAPTON®). The bottom side of the lower film is completely metallized with 1 mil of copper. The top side of the upper film contains a one-half-wavelength circular pattern (Fig. 2), located near its center, with a conductor coupled to it and extending to

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[1] Propogation assumed to be in the TEM mode.
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