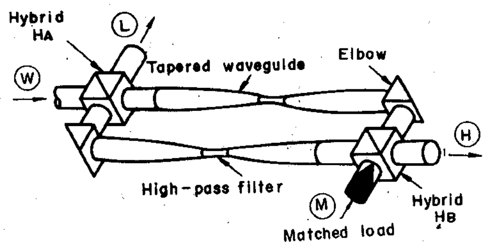

Fig. 1. A Michelson interferometer-type band-splitting filter.

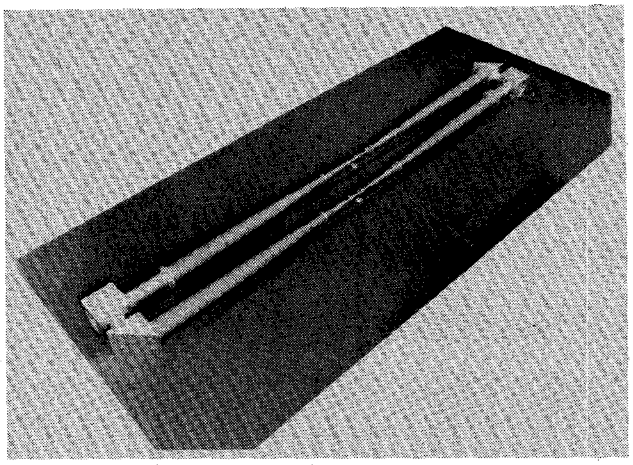

Fig. 2. 40–120-GHz filter.

$f_L - f_C$ will be as small as possible. On the other hand, since the wave in the higher band goes through the hybrids H_A and H_B, the hybrid H_B can be designed such that the overall coupling characteristics of the cascade connection of both hybrids will be 0 dB. In this case, the coupling deviation in the band $f_C - f_H$ must be designed to be as small as possible. Taking this method, we can design the hybrids which double the usable bandwidth as compared to the conventional filters.

This new design method, after all, is to make the coupling characteristics of the hybrids H_A and H_B different from each other: an optimum dielectric constant and thickness of the semitransparent mirror of each hybrid must be properly chosen in a given frequency band. Using the hybrids and cutoff filters, we have constructed a band-splitting filter shown in Fig. 2, whose specifications are as follows:

frequency range	40–120 GHz
cutoff frequency	76.15 GHz
semitransparent mirrors	H_A:F2 glass ($\epsilon_r = 7.2$; $h = 0.41$ mm)
	H_B:potash–soda–glass ($\epsilon_r = 6.8$; $h = 0.27$ mm)
taper profile of cutoff filter	cosine to the third power
waveguide diameter	51 mm
overall length	2.67 m

Experimental results of the filter are shown in Fig. 3. Its branching characteristics are flat. The branching loss is 1.37–2.22 dB in the lower band (*W–L*, 40–76.15 GHz) and 1.16–1.62 dB in the higher band (*W–H*, 76.15–120 GHz). A reflection coefficient in the lower band is below -20 dB (VSWR: below 1.22), residual coupling in the higher band is below -13 dB (below -20 dB in 76.15–110 GHz), and the guard band is as small as 348 MHz.

It is concluded that a Michelson interferometer-type band-splitting filter can be used for a 40–120-GHz millimeter-wave waveguide transmission system.

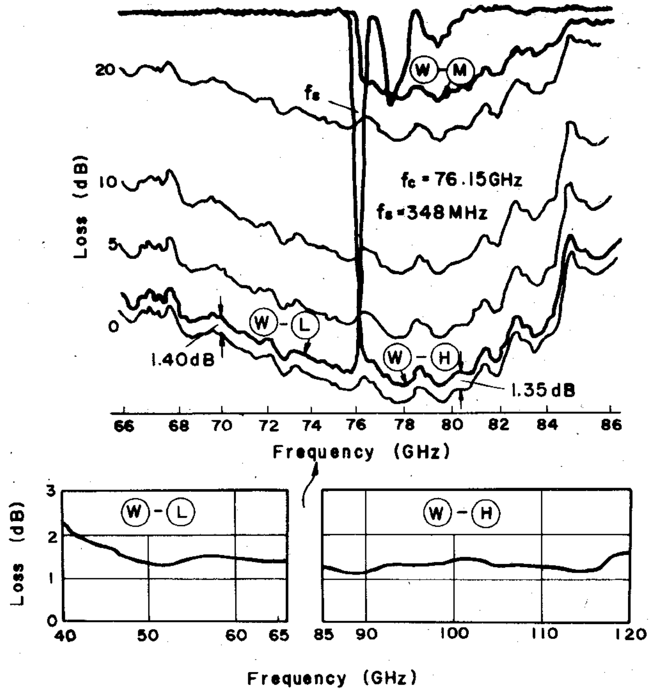

Fig. 3. Measured frequency responses.

ACKNOWLEDGMENT

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Excess Surface Resistance Due to Surface Roughness at 35 GHz

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Abstract—**The increase of the surface resistance of plane copper surfaces caused by mechanically generated surface roughness has been determined at 35 GHz by measuring and evaluating the Q values of an H-guide cavity with removable sidewalls. The sidewalls were ground one-directionally by using abrasive papers of various**

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grades to produce various degrees of roughness. The rms values of roughness were measured mechanically and optically after calibration by microphotography.

INTRODUCTION

Discrepancies exist between measured and computed values of the surface resistance of conductors at high frequencies. The discrepancies were observed when the measured resistance values were compared with those computed from the bulk value of resistivity. They are relatively low at 10 GHz (about 2–5 percent) but become appreciable with increasing frequency. In the millimeter-wave region, ratios of the measured and computed surface resistances of copper up to 1.5 and 2.5 have been observed at 35 and 70 GHz, respectively. Since the frequency range of practical use of radio waves is being extended at present in communications, radar, and other applications into the millimeter-wave region and since reliable values of the surface resistance of conductors are important in design of millimeter-wave components and systems, the discrepancies have become a serious problem.

The discrepancies were observed in device-oriented studies dealing with microwave components, such as waveguides and resonators at S band (2–4 GHz) and X band (8–12 GHz) frequencies [1]–[6]. A few investigations were made in the millimeter-wave region [7]–[9]. The published results do not agree and do not permit definite conclusions. Geometrical surface roughness was usually considered to be the origin, and the increase of surface area due to the roughness was given as an explanation. Since the observed surface effects at frequencies below 15 GHz and the measurement errors were of the same order of magnitude, the results in this frequency range were not conclusive. Similarly, no agreement among authors exists on one specific method such as chemical polishing, electrical polishing, and annealing.

The most recent studies were carried out on waveguides at frequencies in the 35- and 70-GHz regions [8], [9]. Surface roughness and cavity-type irregularities of the surface, caused by the drawing process at the production of the waveguides and observed with a scanning-electron-beam microscope were considered as the origin of the discrepancies. Since measurements in our laboratory on machined and polished copper surfaces which have no such cavities gave similar values of excess resistance as in the above case, the question arises whether geometrical surface roughness and irregularities of the surface may be just one of a larger number of contributions to the excess losses leading to the discrepancies.

For obtaining more conclusive results, a study has been carried out with emphasis on high accuracy of the electrical measurements to be able to separate the various sources contributing to the excess losses and on a satisfactory determination and description of the surface characteristics.

BASIC APPROACH

Plane surfaces with currents flowing in only one direction were chosen to be studied in an H-guide cavity with removable sidewalls. The cavity, shown in Fig. 1, has a rectangular shape with a dielectric slab made of a low-loss ceramic ($\epsilon_r \approx 10$) located in the center. It is operated in the 119 quasi-TE mode (E field parallel to the sidewalls) at 35.2 GHz. Its Q value was measured and evaluated. The fields decrease, typically for H-guide modes, exponentially from the center toward the upper and lower walls. At these walls, the field-strength components are 0.13 of their values in the center. By this, the contributions to the excess losses and to the Q value by the upper and lower walls and by currents crossing the gaps between these walls and the sidewalls are reduced to negligible values. The repeatability of measuring the Q values before and after repeated dismantling of the cavity was about ± 1.5 percent.

The inner surfaces of the sidewalls which carry currents flowing in the vertical direction only were mechanically ground and polished to find the effects of these processes on the surface resistance. Changes of the total Q value were measured and the local Q values of the sidewalls from the overall Q value of the cavity found. The sidewalls were ground to have one-dimensional roughness with the grooves perpendicular to the direction of the current flow. The cavity was used in the transmission-type mode and coupling taken into consideration by evaluation of the insertion loss.

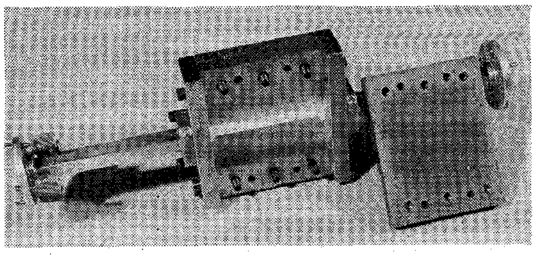

Fig. 1. H-guide resonator with one of the sidewalls removed.

The various contributions to the total unloaded Q value of the cavity can be considered by writing

$$1/Q_{\text{total}} = 1/Q_s + 1/Q_d + 1/Q_d + 1/Q_e + 1/Q_{t\&b}.$$

For a typical case (polished sidewalls), when a total Q value of 9895 was measured, the following figures were found:

sidewalls	$Q_s = 17\,630$
dielectric	$Q_d = 37\,320$
endwalls	$Q_e = 72\,180$
top and bottom walls	$Q_{t\&b} = 269\,900.$

Roughness of the sidewalls affected Q_s only. Its magnitude was found from the measured Q_{total}. The surface resistance is then

$$R_s = G/Q_s$$

where G is a geometrical factor which depends on the structure and the dimensions of the cavity.

ELECTRIC MEASUREMENTS AND SURFACE HANDLING

The Q-value measurements were carried out by slowly sweeping the frequency through resonance of the cavity. The Q value was found on a double trace oscilloscope by evaluating the resonance curve at the intersection with the mode pattern of the klystron on the second trace of the oscilloscope. The half-power bandwidth was obtained by using the interference pattern generated by mixing and frequency doubling signals of various oscillators and superimposing them on the mode-pattern trace. The basic error in determining the Q value was about ± 1 percent.

After trials using other methods, surface roughness was generated by grinding the plane copper surfaces on commercially available grinders using alternatively six grades of abrasive papers under dry and wet conditions. After first carefully polishing the surfaces, grooves were generated by moving the sidewalls back and forth in only one direction applying relatively light force. Grain sizes of the papers varied from 15 to 62 μin according to the manufacturers information. The ground sidewalls and samples were stored in nitrogen to avoid oxidation and nitrogen was blown through the cavity at the Q-value measurements to avoid the effects of oxygen and water vapor.

Determination of the surface roughness was carried out in several ways. A stylus-type profilometer (Bendix Corporation) was used to find the rms value of the roughness mechanically. After an initial comparison of this method with optical evaluation, optical reflectance measurements were used solely. The optical measurements were carried out on an optical-bench setup indicated in Fig. 2(a). The samples under evaluation were mounted on a turntable with the grooves oriented vertically, and the horizontal angular distribution of the intensity of a reflected laser beam was measured as a function of the angle of reflection. The angle of incidence was kept constant at 20°. Diagrams of the angular distributions of the reflected light intensities are shown in Fig. 2(b) for several degrees of surface roughness. The diagrams confirm that the angular spread between specified values of the light intensity normalized with regard to its maximum value is a suitable measure of the surface roughness. The spread

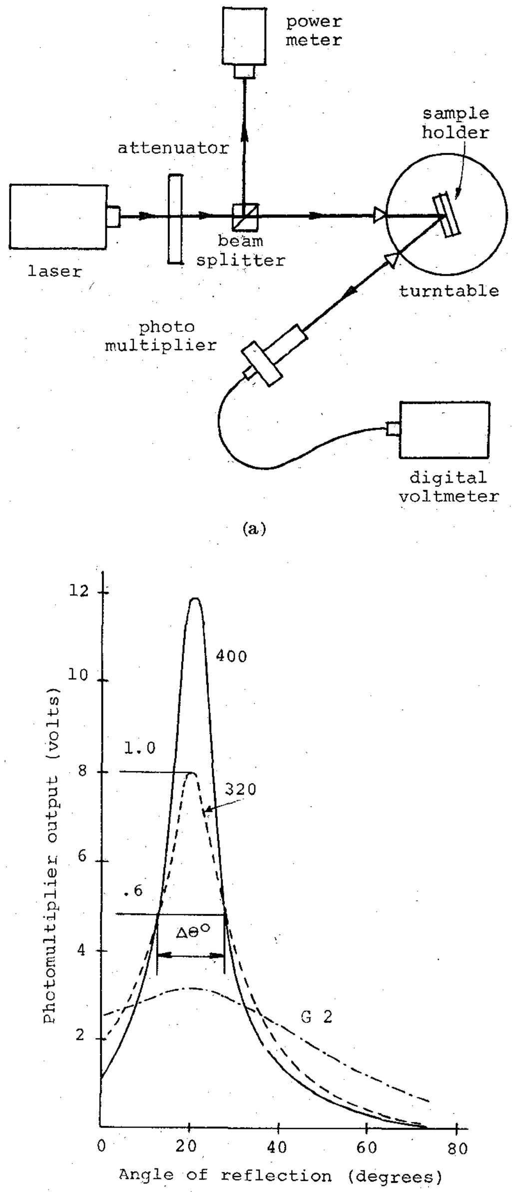

Fig. 2. (a) Optical setup for reflectance measurements. (b) Reflected light intensity.

between the values of 0.6 of the maximum was chosen for the description of the roughness.

Microphotography was then used for the calibration. Herewith a sample surface ground by a specified abrasive paper and evaluated optically was cut and embedded in Bakelite. After grinding and polishing, the boundary line between the ground copper surface and the Bakelite was photographed. One of the Bakelite cylinders containing three ground sample surfaces is shown in Fig. 3(a). The sample surfaces were cut from larger surfaces such as that shown next to the cylinder. An enlarged print showing a typical example of the boundary line can be seen in Fig. 3(b). The figure includes a 10-μm grid for scaling. Prints of such microphotographs of the sample surfaces with a magnification of about 800 were then evaluated. The evaluation gave the rms values, probability distributions of surface levels, and correlation functions. The rms values found in this way were then used for the calibration of the optical measurements. A comparison of the mechanical, microphotographical, and optical data is shown in Table I. The measurements by the profilometer did not agree with the other methods. Optical evaluation of surface roughness and optical description after calibration by microphotography was then used exclusively as a measure of roughness.

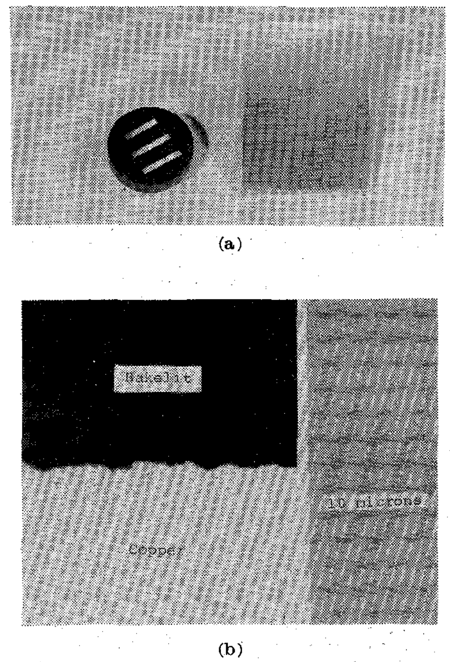

Fig. 3. (a) Bakelite cylinder with three sample surfaces and test surface from which sample surfaces were cut. (b) Enlarged print showing the boundary between a rough copper surface and Bakelite and a 10-μm grid for scaling.

TABLE I
COMPARISON OF MECHANICAL- MICROPHOTOGRAPHICAL- AND OPTICAL-ROUGHNESS DATA

Abrasive paper	Profilometer reading (RMS)	Microphotography (RMS)	Angular spread
G2	0.5–0.7μm	1.30μm	39°
240	0.33–0.38	1.14	20°
320	0.28–0.33	0.96	14°
400	0.10–0.13	0.78	10°

MEASUREMENT RESULTS AND CONCLUSIONS

After studies of the sample surfaces and after calibration of the optical setup by microphotography, several sets of Q-value measurements were carried out using the H-guide cavity. The results are shown in Fig. 4. The diagram shows normalized values of Q_s of the sidewalls as a function of the angular spread of the reflectance as a measure of roughness. The maximum indicated roughness of 40° spread corresponds to an rms value of the roughness of 1.25 μm. Mechanically polished surfaces gave an angular spread of about 2° corresponding to a roughness rms of about 0.5 μm. Both rms values were obtained by microphotography. A normalization was carried out by using the Q_s value for the polished surfaces as 1.0 and by dividing the measured values for various degrees of roughness of the same set by the value for the polished surface. This was done since various sets of surfaces fabricated from slightly different copper material and processed at different times gave slightly different Q_s values when polished. The diagram shows that the maximum roughness associated with the maximum spread reduces the Q value by about 18 percent below the value for polished surfaces. It is interesting to note that the Q_s value for mechanically polished surfaces, which corresponds to a value of about 0.5 μm rms, was found to be about 0.7 of the value obtained by computing Q_s using conventional skin effect calculations and based on the bulk value of conductivity at dc. In the present case, the bulk value was measured to be 5.52×10^7 S/m. This result corresponds to a ratio of measured to computed surface resistance of 1.43. The equivalent surface conductivity would be 0.49 of the above bulk value. After electrolytic polishing,

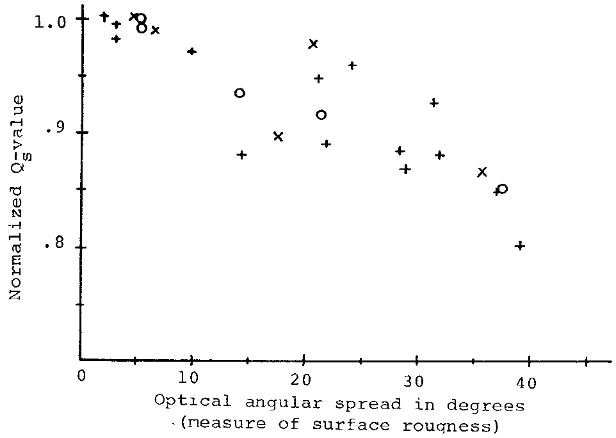

Fig. 4. Normalized Q_s values for the H-guide resonator.

the value of the resistance ratio (1.43 for mechanical polishing) dropped to about 1.3. The errors of the total Q-value measurements including the effects of cavity dismantling were about ± 2 percent. The results clearly show the effect of surface roughness. They also indicate that other effects such as work hardening, oxidation, surface defects, and general surface effects also contribute to excess losses and thus contribute to the discrepancy between measured and computed values of the surface resistance.

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Microwave Amplifier Using Several IMPATT Diodes in Parallel

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Abstract—IMPATT diode amplifiers are described that use several packaged diodes in parallel in a coaxial housing. With a pair of GaAs Schottky-barrier diodes, a power output of 8 W (input locking power equals 300 mW) was obtained at 4 GHz without exceeding safe operating temperatures. Similarly, three-diode circuits produced 15 W (locking power equals 3.5 W) at 4 GHz and >10 W (locking power equals 2.7 W) at 6 GHz under safe operating conditions. The maximum power obtained from the pair was 11 W. The maximum power obtained from the 4 GHz three-diode circuit was 21 W. The efficiency of the diodes at the maximum power level was 12–13 percent. The characteristics of the pair are compared with those of the individual diodes and it is concluded that this power-combining scheme is very efficient and should be economically advantageous.

The scheme permits the total diode area utilized in a single cavity to be increased significantly beyond that which is practical in a single diode package. The use of parallel operation permits efficient heat sinking of each diode package, which is impractical with series operation. The technique employed has been shown to be suitable for extension to three or more diodes for higher power. It is required that each set of diodes be matched for similar I–V characteristics. With this constraint, the close RF coupling of the diodes in conjunction with appropriate stabilizing resistor(s) assures that the diodes operate cooperatively as a unit capable of being powered from a single current-regulated source.

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INTRODUCTION

The need for high-power IMPATT microwave amplifiers imposes the problem of economically combining diodes for maximum power output. Present schemes [1], [2], [5] combine oscillators rather than diodes. They require individual oscillator circuits and separate current-regulated power supplies. One alternative approach is to connect diodes in series in the same resonator circuit [3]. This approach is unrealistic for our projected power levels (>5 W) because of the heat-sinking problem [3]. Parallel connections of two and three diodes are therefore considered in this short paper. The individual diodes are designed as large in area as is practical taking into account thermal, bonding, and RF impedance considerations.

DESIGN CONSIDERATIONS

The purpose of this investigation was primarily to demonstrate the possibility of adding power by paralleling two packaged diodes. Ideally, this requires perfectly matched sets of diodes. This would require matching with respect to their breakdown voltage, susceptance, and negative conductance. These three requirements are not necessarily related.

Preliminary tests showed that a breakdown voltage range ± 1 V was practical and acceptable for the experiment, pending further investigation. No special requirements were imposed on the diode impedance. The oscillator circuit, which was a coaxial circuit for low Q, had to satisfy several requirements.

1) *Diode Mounting*: The diodes had to be mounted such that the coaxial transformer could slide into very close proximity of the diodes to permit tuning of large-area (high-capacitance) units.

2) *Bias Circuit*: The bias should be introduced in such a way as to directly complement the impedance seen by the diode and to avoid uncontrollable impedance transformation out of the operating band of the amplifier.

3) *Amplifier Mounting*: The coaxial housing was to be mounted in the plane of the suspended stripline to permit ready integration into the stripline housing and eventual replacement of the hybrid coaxial-stripline configuration by an all-stripline design. This design would take advantage of the whole amplifier housing as a heat sink.

COAXIAL HOUSING

The coaxial part of the amplifier is shown in Fig. 1. The diodes were arranged in V shape rather than parallel to minimize the series inductance of the circuit and to facilitate a simple connection between the center contact of the diodes and the coaxial center conductor. Bias was introduced in the stripline portion of the amplifier, external to the coaxial housing.

Initial tests of the diode pair were made without the stabilizing chip resistor. This resulted in a nonsymmetrical mode of oscillation