ting above the RF circuit permits the accurate adjustment of the oscillator to a selected frequency before final sealing of the unit and assures reliable performance through repeated temperature cycles.

This transmitter mounted in a projectile shell successfully withstood shock levels of up to 50 000 g and a peak rotational acceleration of approximately 600 000 rad/s^2 during test firing.

ACKNOWLEDGMENT

The thick-film circuit for the modulator–regulator portion of the radiosonde was fabricated by the RCA Defense Electronic Products Division, and the RF transistor chips were mounted by P. McGeough of the RCA Electronic Components Division, both located in Somerville, N. J.

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Recent Experimental Work on Silicon Microstrip Microwave Transmission Lines

In recent years a great deal of literature has become available on microstrip transmission lines, generated by interest in microwave integrated circuitry.[1]–[3] At high frequencies, microstrip on silicon is of interest so that the microwave circuits may be constructed on the same material as the active devices.[4],[5]

Applications involving relatively high power require that the active devices be passivated in order that sufficiently high breakdown voltages be maintained. At present these passivation techniques include the use of grown or deposited silicon-dioxide films, or thick films of glass which may be fired in place where needed.

The effect of these surface coatings on the behavior of silicon microstrip is of interest. New data have been taken on both uncoated and coated structures at frequencies ranging from 3 to 18 GHz. These measurements have been performed on slices with a wide resistivity range and with various surface treatments. Slices nominally 0.010 inch-thick were prepared with a pattern of lines varying from 0.002 inch to 0.030 inch wide. Evaporated aluminum or selectively plated gold conductors of approximately 2 to 3 microns thick were employed.

These data show good agreement with the analytical treatments of Wheeler,[1] and Welch and Pratt.[2] No measurable effects of 10 000 Å grown or deposited SiO_2 were observed on the microstrip transmission characteristics. A 12-micron coating of glass does alter the transmission characteristics of the lines slightly, but little effect on the loss properties was observed.

The slice thicknesses were measured with a hand micrometer using portions of the material which had been cut away from the part with the transmission lines. The line widths were measured with an optical micrometer. Surface-film thickness, line-width thickness, and the glass-layer thickness were measured with a "Tallysurf" micrometer. The error of the width–height data is estimated at ± 3 percent for values on the order of 3. The percentage error grows to ± 15 percent for width–height data $\simeq 0.2$. The resistivity data for the silicon material were taken from Hall-bar measurements.

Fig. 1 presents a plot of experimental characteristic impedance data compared with the analytical curve for silicon. The Z_o data were taken on HP1415A and HP1430 time-domain reflectometers. The data for the unpassivated (uncoated or oxide-coated silicon) slices are from a least-square fit to data of 6 slices. The standard deviation from the fitted curve is 3.1 ohms. The deviation from the analytical curve at low and high characteristic impedances is felt to be a result of the 50-ohm line used for comparison. The data for the glass-coated silicon are from two slices, measured both with the HP1415A reflectometer and the Wiltron 311-310B *Ku*-band impedance analyzer. At *Ku* band, reflection coefficient is displayed on a Smith chart, and corrections[6] are made for line mismatches. Within the range of connector-mismatch and analyzer-unbalance errors, Z_o measurements at *Ku* band generally agree with the time-domain reflectometer data.

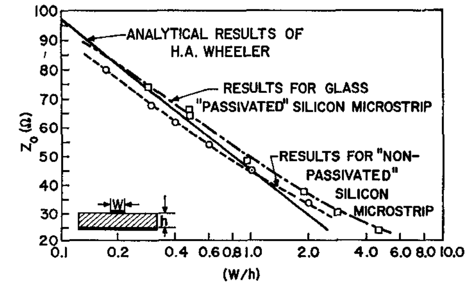

Fig. 1. Characteristic impedance of silicon microstrip.

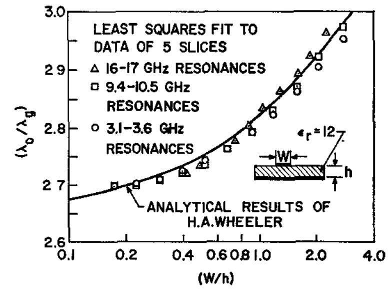

Fig. 2. Slowing factor for plain silicon microstrip for a frequency range of 3 to 18 GHz.

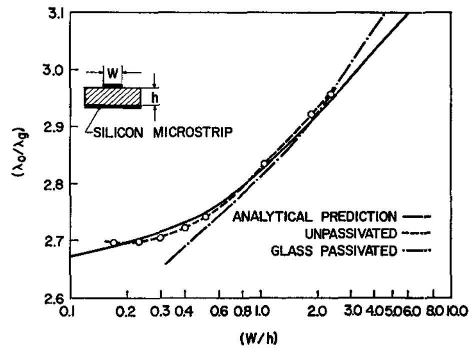

Fig. 3. Slowing factor for glass-passivated silicon microstrip in *Ku* band.

The data shown in Fig. 2 indicate the phase velocity of the silicon microstrip as predicted by Wheeler's analysis for frequencies as high as 18 GHz. Fig. 3 is a comparison of the slowing factor data in *Ku* band for the passivated and unpassivated microstrip. The slowing-factor data (λ_o/λ_g) were taken by measuring multiple half-wavelength resonances from 3.1 to 18 GHz. A slice was capacitively coupled from the generator and to a detector, padded by 10 dB, in such a way that the transmitted signal at resonance was at least 40 dB down from the generator signal. The parallel lines of the slice, other than those being measured, were shorted so that mutual coupling effects could be minimized. The resonance frequencies were measured to ± 1 percent with the use of the marker generators in the

Manuscript received February 16, 1968; revised March 20, 1968. This work was partially supported by the Air Force Avionics Laboratory, Wright-Patterson AFB, Ohio, under Contract F 33 615-67-C-1817.
This special issue on Microwave Integrated Circuits is published jointly with IEEE TRANSACTIONS ON ELECTRON DEVICES, July 1968, and the IEEE JOURNAL OF SOLID-STATE CIRCUITS, June 1968.

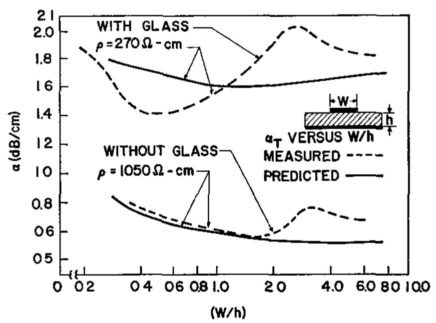

Fig. 4. Loss of plain and glass-passivated silicon microstrip.

lower frequency ranges, and to ±0.1 percent with a cavity wavemeter in *Ku* band. Computer simulation of the circuit, with fringing and coupling capacitors, shows resonant frequency shifts on the order of 1/2 to 1 percent downward at the higher resonant frequencies.

The loss measurements were made by taking the 3-dB point frequencies on the 15- to 18-GHz half-wavelength resonance of the transmission lines. Because the signal level is quite low, and because it is difficult to pick these frequencies exactly, it is felt that the loss measurements should be within 30 percent of the loss value in dB. The loss figures were calculated as follows:

$$Q = \frac{f_0}{\Delta f} = \frac{8.686\beta_0}{2\alpha(\text{dB/cm})} \tag{1}$$

and

$$\alpha\left(\frac{\text{dB}}{\text{cm}}\right) = 0.9102\cdot\Delta f(\text{GHz})\cdot(\sqrt{K^\Phi}) \tag{2}$$

where

β_0 = phase constant at resonance
f_0 = resonant frequency
Δf = 3-dB bandwidth
$8.686 = 20 \log_{10} e$
K^Φ = effective dielectric constant ($\sqrt{K^\Phi}$ = slowing factor).

These data are shown in Fig. 4 for slices of 270-Ω·cm glass-passivated silicon and 1050-Ω·cm unpassivated silicon. The losses of silicon microstrip have been tied to Wheeler's filling factor by Welch and Pratt.[2] For metal thicknesses greater than 1.5 skin depths, the conductor losses may be easily approximated, as shown in Caulton, Hughes, and Sobel.[3]

The data in Fig. 4 are in four parts, and show the effect on loss of glass-passivating silicon microstrip. Two slices of the same resistivity were not readily available. The measured attenuation constant on the 1050-Ω·cm material is compared with the predicted curve to establish measurement accuracy. The measured loss for the 270-Ω·cm material is compared with a curve predicted in the same fashion as the previous case, i.e., no provision is made for the glass layer in the predicted curve. The difference in the relative comparison of measured to predicted shows there is no effect of glass passivation on the loss. The data agree with Hyltin's[4] loss data. Both measured curves show an inflection near width–height ratio = 2. The data in both cases were somewhat erratic but more so in the passivated case. The dotted lines representing measured data are "best fit" curves only, and show the average of a spread of data points.

In conclusion, we see that near-TEM transmission occurs for silicon microstrip of convenient thickness for frequencies up to *Ku* band. The loss of silicon microstrip is predicted within the present experimental error by the existing theory. Addition of a second dielectric layer of small thickness slightly affects the propagation characteristics, but does not affect the loss.

Acknowledgment

The authors gratefully acknowledge the assistance of S. Emmons, B. Hooper, and H. Leedy for their help in preparing slices. Microwave measurements were performed with the assistance of B. Myers and G. Hollis.

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A Microstrip Nonreciprocal Tunable YIG Filter

***Abstract*—A nonreciprocal tunable YIG filter in a microstrip configuration has been constructed which makes use of a novel method of generating a circularly polarized field in the plane of a microstrip circuit. Nonreciprocities in excess of 40 dB have been obtained at *X* band with relatively low insertion loss and VSWR.**

This correspondence presents some preliminary results obtained on nonreciprocal tunable YIG filters designed in a microstrip configuration for use in integrated microwave circuits. Nonreciprocal filters of this type would find application as tunable resonance isolators, tunable band-reject filters, and, with proper coupler design, tunable or electronically swept nonreciprocal bandpass filters. The use of a unique method for generating a region of circular polarization in the plane of a microstrip circuit has led to the development of nonreciprocal band-reject filters with characteristics similar to those obtained with waveguide devices.[1],[2] Previous investigations of circular-mode generation involved distorting a TEM mode to a TE mode by using an inhomogenous dielectric. Fleri and Hanley,[3] as well as Anderson and Hines,[4] reported good isolation ratios for dielectrically loaded striplines. However, the size and/or fabrication problems associated with loaded structures precluded their compatible use in integrated microwave circuits. The small size of a YIG sphere requires a circularly polarized field to exist only over a relatively small region in a microstrip structure. To this end, a number of methods for generating circular polarization at a point in a microstrip circuit were investigated,[6] one of which is presented here.

The design and operation of the filter is relatively straightforward, as can be seen from the accompanying figures. Its operation is based on a field-shaping technique which will provide a circularly polarized "*H*" field at a point in the transmission line which is in the vicinity of the YIG material. This will provide a method of nonreciprocally coupling to the YIG; the amount of this coupling is dependent upon the sense of the RF circular polarization relative to the sense of the electron precessional rotation in the material. The method of circular-field generation is readily understood when the requirements neces-

Manuscript received January 29, 1968; revised March 11, 1968.
This special issue on Microwave Integrated Circuits is published jointly with IEEE Transactions on Electron Devices, July 1968, and the IEEE Journal of Solid-State Circuits, June 1968.