TABLE I
DESIGN EXAMPLE

Band edge $f_1 = 11.9$ $f_2 = 12.1$ (GHz)	
VSWR at center frequency 1.5, $g = 0.57$	
Waveguide 19.05×9.525 (mm)	
Sheet thickness 0.5 (mm)	
$d_1 = d_3 = 2.976$ $d_2 = 8.554$	
$L_{12} = 12.273$ (mm)	

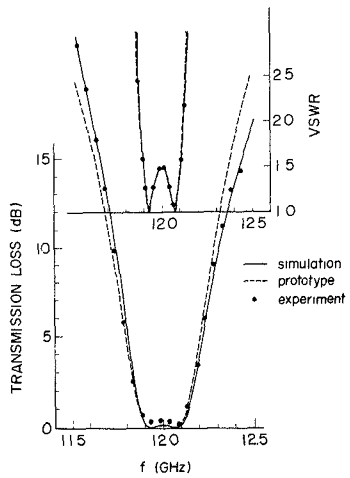

Fig. 5. Designed filter characteristics.

converted to the equivalent circuit shown in Fig. 1(b). Filter design can be deduced from the design technique described in a "synchronously tuned reactance coupled filter" [3]. Applying the technique, step reactances, $u_i = B_i/Y_0$ $(i = 1,n + 1)$, are determined from the desired filter characteristics (Chebyshev, maximally flat). The strip length d_i that has a necessary value of u_i can be found from Fig. 3. Separation between the strips θ_i is given by

$$\theta_i = \psi_i - \Phi_i + \psi_{i+1} - \Phi_{i+1} + \frac{\pi}{2} \tag{10}$$

where

$$\psi_i = \tfrac{1}{2}\tan^{-1}\left(\frac{u_i}{2}\right).$$

Here, Φ_i for necessary d_i is given by Fig. 4.

θ_i is frequency dependent and its dependency is greater than the quarter-wave prototype transformer by a factor of

$$\frac{1}{\beta_i} = \frac{2}{\pi}\left[\frac{d\theta_i}{d(\lambda g_0/\lambda g)}\right]_{\lambda g = \lambda g_0} = g_i + g_{i+1} + 1 \tag{11}$$

$$g_i \equiv \frac{2}{\pi}\left[\psi_i - \Phi_i - \frac{d(\psi_i - \Phi_i)}{d(\lambda g_0/\lambda g)}\right].$$

Consequently, the bandwidth of the filter w will be smaller, compared to that of a quarter-wave prototype transformer w_q, and

$$w = \beta w_q. \tag{12}$$

This contraction of the bandwidth must be considered when calculating the prototype bandwidth w_q from desired characteristics. Contraction factor β in (12) should be the smallest among β_i $(i = 1 \cdots n)$ in (11). A value of g is calculated as about 0.57 at 12 GHz giving $\beta = 0.467$.

Table I and Fig. 5 show an example of a filter design. The dotted line in Fig. 5 shows the characteristics of a stepped half-wave filter prototype determined by the input data in Table I, while the solid line shows a simulation of the designed filter. Both of them have almost the same characteristics at the passband, though their skirt cutoff responses differ slightly from each other. Simulation indicates less attenuation at the upper band edge and more attenuation at the lower band edge. This is mainly due to the frequency dependence of u, as is shown in Fig. 3.

The filter was made with the dimensions in Table I. The dimensional accuracy was necessary only for the center conductive sheet with holes. We made it by the arc discharge machinery, but alternately this could be done by the die for the mass production.

In Fig. 5, experimental data are also illustrated by dots, which show good agreement with the simulation. This indicates that the constant current assumption we used during the analysis was good enough for practical use.

V. CONCLUSION

The waveguide-sandwich filter originally proposed by Konishi has an advantage in its structural simplicity. The inductive strips with finite thickness, which are the basic elements of the filter, were analyzed to establish an equivalent circuit. Charts for the filter design were proposed, and a filter was designed for an M-band waveguide. The experiment was in good agreement with the simulation of the designed filter.

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Conduction and Radiation Losses in Microstrip

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***Abstract*—Losses in microstrip on fused-silica and alumina substrates have been experimentally evaluated for various values of stripwidth. Radiation losses in circuits, commonly used for measuring losses in microstrips, depend on the dimensions of the circuit and affect the total losses considerably. Radiation and conduction losses of open-ended line resonators have been separately determined. These measurements have also drawn attention to discrepancies between published theories and our experiments. The effect of surface roughness upon conduction losses has been measured.**

Many measurements of attenuation of microstrip on alumina substrates have been reported previously. In this short paper data will also be presented for microstrip on fused silica. The relevant properties of the fused-silica and alumina substrates are summarized in Table I. The alumina substrates are selected to meet these specifications and are, in most cases, anisotropic with the highest dielectric constant perpendicular to the surface [1]. This latter value is given. The surface roughness of the silica substrates can be easily controlled, providing the possibility to investigate the relation with conduction losses. The conductors consist of two layers. A nickel–phosphorous adhesion layer is deposited by an electroless plating

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TABLE I

RELEVANT PROPERTIES OF SUBSTRATES USED FOR THE EXPERIMENTS

SUBSTRATE	FUSED SILICA	ALUMINA
THICKNESS	0.5 ± 0.005	0.635 ± 0.01 (mm)
DIEL CONST	3.78	10.8 ± 0.1
ROUGHNESS	0 – 0.7	0.2 – 0.4 (μm)
CONDUCTOR LAYERS	0.1 μm Ni-P, 7 μm Au	

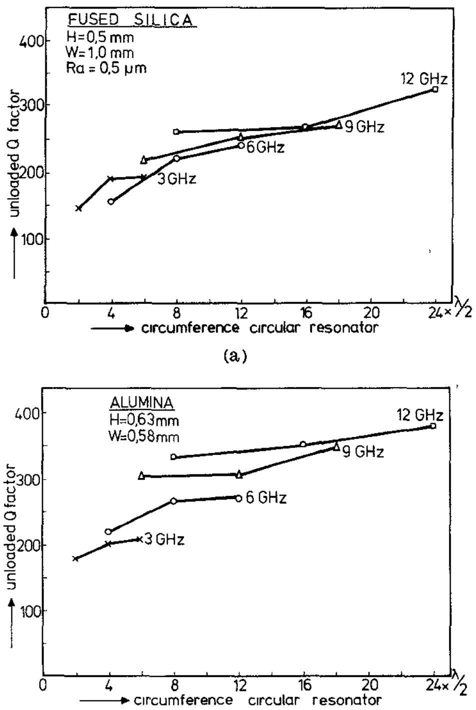

Fig. 1. Q factor of three different ring resonators with a line impedance of 50 Ω. H and W denote the substrate thickness and stripwidth, respectively. (a) On fused silica. (b) On alumina.

process, with a 7-μm gold layer on top of it. Low line losses are easily determined with a resonance measurement. The most commonly used resonator circuits are the ring resonator [2], [3] and the shielded and unshielded open-ended line resonator [1], [4], [5].

The unloaded Q factors of three ring resonators with different diameters on alumina and fused-quartz substrates are presented in Fig. 1(a) and (b).[1] Parasitic coupling between the input and output ports via surface waves was prevented by placing absorbing material on the substrate. It is obvious (from Fig. 1) that if the radius is of the order of a few wavelengths, the loss increases with increasing curvature of the line. This is most likely due to the excitation (by the curved line) of surface waves, that are dissipated in the absorbing material on the substrate. At 12 GHz (and already at 9 GHz for alumina) another effect occurs, probably caused by radiation as can be concluded from the measurements to be described. From these results we may conclude that the ring resonator is less suitable for accurate measurements of microstrip losses because these losses are affected by the ring diameter.

Unshielded open-ended line resonators exhibit excessive loss at the open ends because of radiation. By comparing the values of the unloaded Q factors of unshielded resonators and similar ones in a waveguide below cutoff, radiation losses can be determined, while the effective dielectric constant for the pseudo-TEM mode is almost unperturbed [1], [5]. The conduction losses in the shielding can be neglected. The unloaded Q factors of the shielded and unshielded microstrip resonators with a characteristic impedance of 50 Ω on fused silica with different surface finishes are plotted in Fig. 2, where Ra denotes the surface roughness.[1] The curve derived from Pucel's formulas is also presented [10].[2] In these calculations the conductivity of the gold layer is assumed to be the same as that of bulk material. Mostly, a plated gold layer possesses a lower conductivity. This set of curves reveals two interesting aspects.

Firstly, the relationship between radiation and frequency is more complex than suggested in the literature, where *only* the open ends of the line are assumed to excite radiation [5], [6]. In fact, the whole resonant line acts like an antenna with maximum radiation loss if the line length is approximately equal to an integral multiple of half-wavelengths in free space. The higher the frequency, the higher the radiation loss due to the proximity of the ground plane. Consequently, maxima and minima will occur in the radiation and also in the Q factor. These measurements reveal that the "antenna effect" of unshielded microstrip resonators produces a substantial part of

[1] Ra denotes the mean square root value of the surface roughness, measured with a "Talysurf 4."

[2] The calculated losses include the dielectric losses assuming for silica and alumina a loss tangent of, respectively, 10^{-4} and 2.10^{-4} at all frequencies.

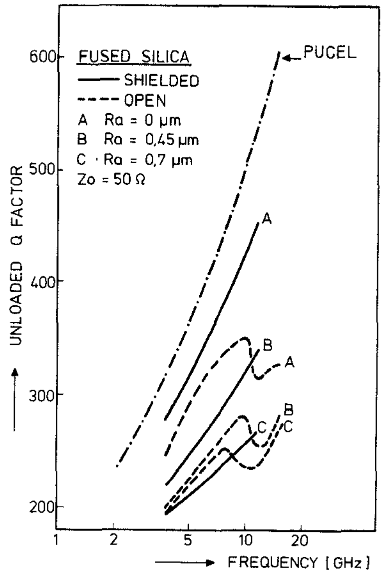

Fig. 2. Q factor of open-ended line resonators (shielded and open) on fused silica substrates with various surface roughnesses.

the radiation losses. It is obvious that the radiation excited at the open ends and by the antenna effect cannot be separated. The unshielded Q factor cannot be measured accurately. The resonance curve is asymmetrical. Similar radiation losses occur at higher frequencies with the ring resonators, as already mentioned before.

The second effect in Fig. 2 is the degradation of the line quality with increasing surface roughness. The Q factors of the line on polished silica agree rather well with the calculated values. Similar results have been obtained on alumina substrates. We may conclude that the shielded open-ended line resonator in a waveguide below cutoff gives reliable values of conduction losses in microstrips, and that this circuit is essentially better than circuits in open test jigs. This is particularly important when evaluating different technologies and aging effects in circuits. In practical circuits shielding below cutoff will not always be possible. Parasitic coupling and radiation then will occur, but depend largely on the structure and the enclosure.

We have also found that radiation does not increase significantly with decreasing dielectric constant for the values 4 and 10. The following data have been calculated from [5] and [6] for radiation losses of 50-Ω lines:

P_r/P_{dist}	According to [6] at 1 GHz (percent)	According to [6] at 10 GHz (percent)	According to [5] at 1 GHz (percent)	According to [5] at 10 GHz (percent)
silica	0.4	85	0.8	50
alumina	0.4	75	0.5	30

From our measurements we found for silica and alumina at 10 GHz approximately 30 and 25 percent, respectively. However, due to the complicated relation between frequency, line length, and radiation loss, no firm conclusion can be made about the validity of the theory in [5] and [6].

Resonators on alumina and fused silica with line impedances of 45, 50, 80, and 100 Ω have been measured. For each line impedance several samples were used. The reproducibility of the data obtained with each circuit was better than 1 percent. However, the maximum difference between the samples was found to be 10 percent using the same technology and line impedance. In Fig. 3 two typical examples are presented together with the calculated curves for fused silica. The best fitting straight line was drawn through the measured points. The slope of this line differs slightly from the calculated one because of the surface roughness of the substrate. From this line, the values are derived at other frequencies.

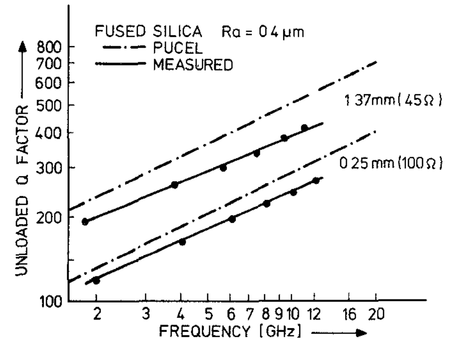

Fig. 3. Typical values of measured Q factors of shielded open-ended line resonators on fused silica, compared with calculated curves.

In the top portion of Table II typical values of the Q factors are summarized for four lines with various impedances together with the percentage of Q measured over Q calculated. At higher frequencies this percentage decreases due to surface roughness. At lower impedances the percentage decreases, which is possibly due to transversal currents. For narrow lines, having a higher impedance, these transversal currents are considerably smaller. Similar effects can be observed for lines on alumina in the bottom portion of Table II. It is interesting to note that for higher impedances the measured Q factor exceeds the calculated value. This can be partly explained by the anisotropy in the alumina and partly because the calculations are based upon the assumptions of a TEM mode. In practice one would expect a lower Q value than the predicted one, due to surface roughness and a lower conductivity of the deposited gold layers than the value of bulk material as used in the calculations. Irregularities along the edge of the line will also increase the losses.

In Fig. 4 the measured increase in attenuation of rough substrates with respect to polished substrates is compared with curves given by Morgan [8] and Lending [9].[3] Morgan calculated the increase of losses for three different types of grooves perpendicular to the current flow. The agreement with the results of Morgan is surprisingly good, taking into account the rather poor accuracy of the roughness measurements and the simplifications in the theory.

[3] The Q factors of resonators on alumina are compared with resonators on "super smooth" alumina ($Ra \leq 0.1$ µm).

TABLE II

TYPICAL VALUES OF MEASURED Q FACTOR OF SHIELDED OPEN-ENDED LINE RESONATORS, WITH THE RATIO Q MEASURED OVER Q CALCULATED

FUSED SILICA Ra = 0.4 µm

FREQ [GHz]	45 Ω Q_0	45 Ω %	50 Ω Q_0	50 Ω %	80 Ω Q_0	80 Ω %	100 Ω Q_0	100 Ω %
2	210	85	200	91	160	97	120	92
4	265	81	265	83	210	92	165	89
6	310	78	305	79	250	89	195	86
8	350	77	340	77	280	86	220	84
10	385	75	370	75	310	86	240	82
12	415	74	390	73	340	85	260	82
14	440	73	415	71	360	84	280	82

ALUMINA Ra = 0.2 µm

FREQ [GHz]	45 Ω Q_0	45 Ω %	50 Ω Q_0	50 Ω %	80 Ω Q_0	80 Ω %	100 Ω Q_0	100 Ω %
2	220	96	190	88	125	100	80	103
4	300	97	265	89	175	100	110	100
6	370	97	320	89	215	100	130	100
8	420	98	370	90	245	100	150	100
10	470	98	410	90	270	100	165	97
12	510	98	440	90	300	100	180	97
14	550	98	470	90	320	100	195	97

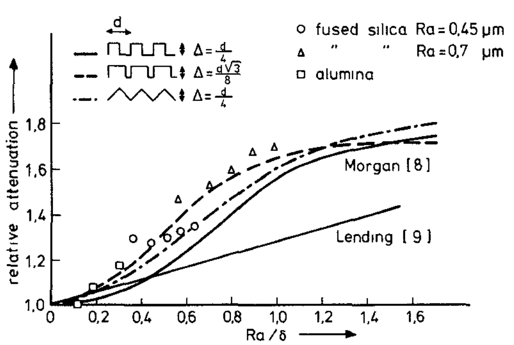

Fig. 4. Relationship between surface roughness and increase of loss in shielded open-ended line resonators on fused quartz. The skin depth is denoted by δ.

From our experiments we may conclude the following.

1) The most meaningful and accurate method of evaluating conduction losses in microstrip is the resonance method with a resonant circuit in a conducting enclosure, preventing other propagation modes but the wanted one. An open-ended line resonator in a waveguide below cutoff is a very useful solution.

2) Pucel's formulas provide a good approximation. For higher impedances the actual losses seem to be somewhat lower than the predicted ones.

3) The curves of Morgan agree well with our experiments and can be used as a first approximation.

4) Radiation losses of microstrips are not so simply related to the frequency as has been suggested. The layout and the surrounding objects affect the radiation losses significantly.

5) Low- and high-dielectric-constant materials show approximately equal radiation losses.

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