Narrow-Band Microstrip Bandpass Filters with Low Radiation Losses for Millimeter-Wave Applications

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Abstract —Microstrip lines are attractive for the lower millimeter-wave ranges, but use of relatively thick substrates would be desirable in order to minimize losses. On such substrates the usual types of microstrip narrow-band bandpass filters (formed from, e.g., coupled line segments with open ends) tend to radiate strongly, giving very poor performance. It has been found that a grating technique initially developed for use with dielectric waveguide can be adapted for microstrip to obtain narrow-band millimeter-wave microstrip filters with little radiation and strong filter characteristics. The stopbands are broad, the second passband occurring at three times the frequency of the first passband. These filters use parallel-coupled gratings with a single grating in cascade at each end. In this paper, we detail the modifications to the dielectric waveguide filter theory which are necessary for use with microstrip. We also present experimental results from microstrip realizations which demonstrate their potential for mm-wave microstrip applications.

I. Introduction

THE UNDESIRABLE effects of radiation from sharp discontinuities (open ends in particular) in conventional microstrip filters are well recognized and tend to be the primary cause of losses in narrow-band microstrip filters in the millimeter-wave range [1], [2]. For example, a plot in [2] indicates that at 30 GHz, half-wavelength microstrip resonators on 10-mil-thick Teflon may have their Q's reduced by a factor of more than 10 from what their Q's would be without radiation. Also, using equations in [3] we calculate that for half-wavelength resonators on 11-mil Duriod with $\epsilon_r = 2.2$, the resonator Q's including only radiation loss would be 34 at 30 GHz. In principle, radiation can be suppressed at a given frequency by using thin enough substrates, but that has the effect of increasing the resistive losses in the microstrip lines, which become particularly serious at millimeter-wave frequencies, especially for narrow-band filters. Shielding can also be used around the circuit to help reduce radiation, but that may not always be convenient, and use of a housing may deteriorate the upper stopbands of a filter due to waveguide modes. It therefore seems that a narrow-band bandpass filter configuration that does not suffer from radiation Q-degradation would be useful.

Designing microstrip resonators with thick substrates poses problems similar to those for realizing resonators in dielectric waveguide where radiation tends to occur at all discontinuities. One solution to that problem in dielectric waveguide is the use of parallel-coupled grating filters [4], [5]. Using those kinds of structures it is possible to realize narrow-band bandpass filters in, e.g., image guide, where the energy is much more loosely bound than in a microstrip. Therefore, that filter principle should also work with low radiation losses if realized in microstrip. However, the parallel-coupled-grating principle requires that the coupling between the gratings be of the so-called forward-coupling type, while coupled microstrips exhibit both forward and backward coupling. For this microstrip application the forward coupling, i.e., the coupling due to different odd- and even-mode velocities, is the one that is useful, while the backward coupling, due to the different odd- and even-mode impedances of coupled lines, is undesirable. Design using forward coupling in microstrip is very unusual since such coupling is generally believed to be very weak or even unusable. However, we investigated the possibilities of using the coupled-grating principle with microstrip, and it was found that the forward coupling is strong enough to permit these filters to be realized and that good filter properties can be achieved despite the presence of backward coupling.

The main advantage of these filters is that they have negligible radiation on even thick substrates, so they approach the full potential Q of microstrip resonators, leaving mainly the dielectric and ohmic losses. These kinds of filters also have broad and relatively strong stopbands (the first spurious passband occurs at three times the passband frequency), and the attenuation is absorptive over most of each stopband. The major disadvantage of the type of filters discussed herein is that they are large measured in wavelengths, which, however, may be acceptable in the millimeter-wave range. The passband width of these filters is also limited to a few percent at maximum. On the other hand, for wide bandwidths higher losses can be toleranted so conventional techniques may prove usable.

It is the purpose of this paper to report on our theoretical and experimental findings concerning the adaptation of

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this filter principle for use with microstrip. Since much of the theory for the design of microstrip grating filters is very similar to that for dielectric waveguide filters, we will rely on reference [5] for most of the necessary information. However, there are some important differences between the dielectric waveguide and microstrip cases, and equations will be presented to treat these situations. Also, we present an improved structure for the case of filters with three or more resonators. This structure gives added stopband attenuation over that discussed in [5]. Besides microstrip, it should also be possible to apply these principles to filters designed using other forms of transmission lines, e.g., coplanar waveguide, where radiation at discontinuities may be troublesome.

II. Bandpass Filter Configuration Using Microstrip Gratings

For the purposes of this discussion, the word "grating" is used to describe a periodic arrangement of microstrip-line segments of alternating impedance connected in cascade (see Fig. 1(a)). The line segments are made to have equal electrical lengths, and an equivalent circuit, shown in Fig. 1(b), is used. It is assumed that the impedance steps are so small that fringing effects at the discontinuities can be ignored. However, as explained in [4], the equivalent circuit shown in Fig. 1(b) actually has a wider range of applicability and could be used even if the line segments were not electrically equal in length and fringing effects at the steps were taken into account. Gratings such as these (or other grating configurations) behave as bandstop filters, as is well known from the theory of periodic structures. For the circuit shown in Fig. 1(b) the stopband center frequency is the frequency for which all the line segments are a quarter-wavelength long. The grating impedance ratio r, which is an important design parameter, is simply the ratio of the impedances corresponding to microstrip lines of widths W_1 and W_0 (Fig. 1(a)):

$$r = \frac{Z_1}{Z_0} > 1. \tag{1}$$

For bandpass filter applications, pairs of parallel-coupled gratings, as shown in Fig. 2(a), are used in the filter structure along with single gratings. Fig. 3 shows a sketch of a two-resonator grating filter, where the parallel-coupled gratings extend off to the right and have distributed loads (not shown) to make the gratings appear approximately as though they were infinitely long. The gratings are all resonant at the center frequency f_0 of the filter passband. Gratings G_1 and G_{A1} are separated by a multiple of a half guide wavelength, and a bandpass resonance occurs due to reflections between these two gratings. Similarly, gratings G_2 and G_{A2} with the line in between form a second resonator. The length of grating G_1 controls the coupling between the first resonator and the input line, while grating G_2 serves a similar function for the output line. The spacing between the gratings G_{A1} and G_{A2} along with other parameters of these gratings controls the coupling between the resonators. When the gratings are reflecting (i.e., in their stopband), the structure in Fig. 3 functions as a two-resonator reactive bandpass filter. However, when the gratings are not reflecting, the power entering at the upper left propagates to the right along gratings G_{A1} and G_{A2} to the distributed loads off to the right. Since the parallel-coupled gratings are designed to exhibit predominantly forward coupling, very little power is coupled to the output on the left, and high absorptive attenuation is exhibited between the input and output ports on the left. In the case of microstrip parallel-coupled gratings we found it convenient to use notches only on the outer sides of the lines so that the spacing between lines could be kept constant. In order to obtain low radiation, we found it essential to use gradual bends in the lines, as indicated in Fig. 3.

More detailed discussion of the functioning of coupled gratings will be found in [4] and [5]. For analysis of microstrip gratings the equivalent circuit in Fig. 2(b) is used where, for purposes of deriving design equations, the

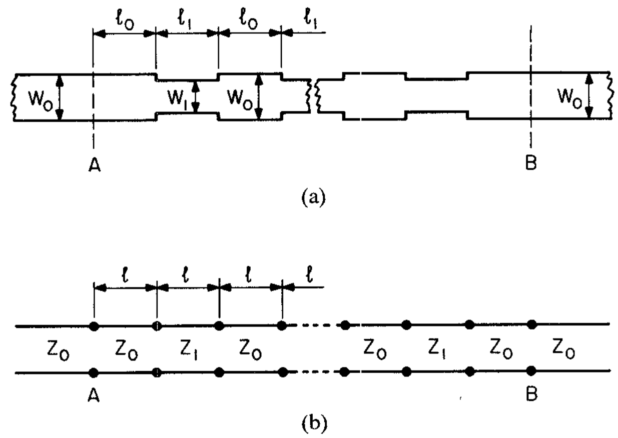

Fig. 1. (a) The strip pattern of a microstrip grating and (b) its equivalent circuit.

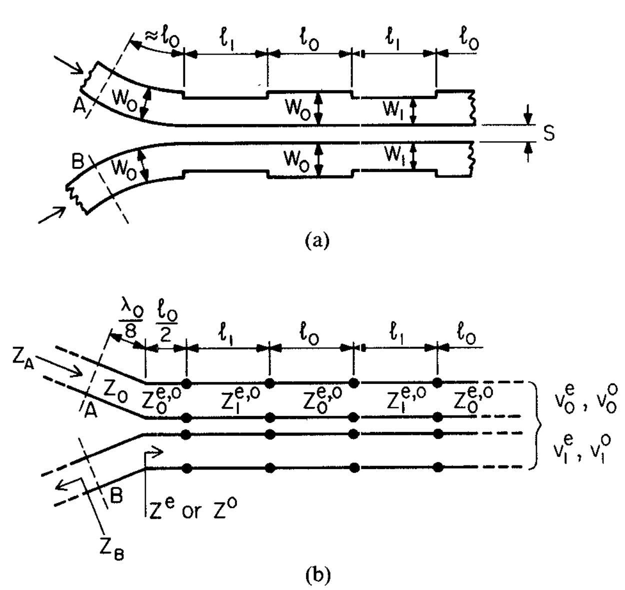

Fig. 2. (a) The strip pattern of a coupled microstrip gratings and (b) the corresponding equivalent circuit.

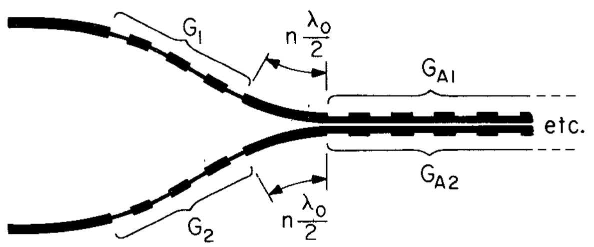

Fig. 3. The strip pattern of a two-resonator filter. The notch depth of the gratings has been greatly exaggerated. The coupled gratings continue to the right; only part is shown here.

structure is assumed to extend to infinity on the right. It should be noted that this equivalent circuit differs from that used in [4] and [5] in that here the l_0 and l_1 line sections have different odd- and even-mode impedances in addition to having different odd- and even-mode velocities.

III. Coupled-Resonator Model of Gratings

For filter design purposes it is convenient to be able to represent the circuits shown in Figs. 1(b) and 2(b) with the two-resonator equivalent circuit shown in Fig. 4. The circuit of Fig. 4 models the circuit in Fig. 1(b) between the reference planes A and B at and near the stopband center frequency of the grating. It also models the circuit in Fig. 2(b) between the reference planes A and B at and near the passband center frequency of the coupled gratings. It is characterized in terms of the resonant frequency f_0, reactance slope parameter x of the resonators, and impedance inverter parameter K_{AB}.[1] This coupled-resonator model is used to synthesize the filter so that it will have a prescribed passband shape and width.

For single gratings the equations developed for the dielectric waveguide case [5] can be used directly because the same equal-length equivalent circuit, Fig. 1(b), applies. For microstrip coupled gratings the theory of [4] and [5] is not directly applicable because it omits the difference between the odd- and even-mode impedances, as was noted above in connection with Fig. 2(b). Generalized equations which can be used with microstrip coupled gratings are developed below.

The method of odd and even modes is convenient for the analysis of coupled gratings. It can be shown that the two-port transfer characteristics of the circuit in Fig. 2(b) can be completely characterized in terms of the impedances Z^o and Z^e that one sees looking into one of the gratings under odd- and even-mode excitation conditions, respectively [4]. For a theoretical analysis it is convenient to assume that the coupled gratings are infinitely long. Methods to calculate the impedance seen looking into an infinite periodic structure are well known. The expressions for these impedances take on especially simple and useful forms if in the equivalent circuit of Fig. 2(b) l_0 and l_1

[1]For a general discussion and definitions of resonator reactance slope parameters and impedance inverters, see [6, chs. 4 and 8].

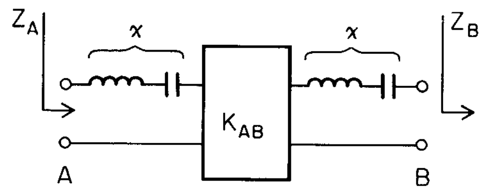

Fig. 4. A coupled-resonator equivalent circuit which is used to model the circuits in Figs. 1(b) and 2(b) at frequencies close to f_0

have equal electrical lengths and if the reference plane for these impedances is set in the middle of the first segment. Then (c.f. [4, eq. (5a)])

$$Z^{e,o} = Z_0^{e,o}\sqrt{\frac{(1+r^{e,o})\cos\Theta^{e,o} - (1-r^{e,o})}{(1+r^{e,o})\cos\Theta^{e,o} + (1-r^{e,o})}} \tag{2a}$$

where

$$\Theta^{e,o} = \frac{\omega}{\omega_0^{e,o}}\frac{\pi}{2}. \tag{2b}$$

The $\omega_0^{e,o}$ are the stopband center frequencies, and the $r^{e,o}$ are the grating impedance ratios; superscript e implies the even mode, and superscript o implies the odd mode. We have found it convenient to choose l_0 and l_1 in Fig. 2 so that their electrical lengths are exactly equal in the odd mode. Then the odd-mode stopband center frequency ω_0^o is simply the frequency for which the odd-mode electrical lengths of l_0 and l_1 are equal to a quarter wavelength, and (2a) can be used straightforwardly to find Z^o. Then, in the most general case, the even-mode electrical lengths of l_0 and l_1 may not be exactly equal. However, for all practical cases the deviation is very small and (2a), with the appropriate even-mode parameters, can be used as an excellent approximation for Z^e. The even-mode stopband center frequency ω_0^e is fixed by the fact that the period of the even-mode grating is a half wavelength. It can be shown that

$$\Theta^e = \left(\frac{v^o}{v^e}\right)_{\text{ave}}\Theta^o \tag{3}$$

where $(v^o/v^e)_{\text{ave}}$ is the average of the odd- to even-mode velocity ratios of the Z_0 and Z_1 segments. The impedance ratios to be used in (2a) (and other appropriate expressions) are

$$r^{e,o} = \frac{Z_1^{e,o}}{Z_0^{e,o}} > 1. \tag{4}$$

As explained in [4] and [5], the coupled gratings behave as a lossless, reactive circuit between the ports A and B in Fig. 2 only at frequencies at which both the odd and even modes for the gratings are simultaneously in their respective stopbands, i.e., where the grating stopbands overlap. (The coupled gratings will exhibit a passband centered within this overlap band [4], [5].) For microstrip gratings the ratio of the edge frequencies f_U (upper edge) and f_L

(lower edge) of this overlap band becomes

$$\frac{f_U}{f_L}=\frac{\cos^{-1}\left(\dfrac{1-r^e}{1+r^e}\right)}{\left(\dfrac{v^o}{v^e}\right)_{\text{ave}}\cos^{-1}\left(\dfrac{r^o-1}{r^o+1}\right)}. \tag{5}$$

Conditions for fixing the center frequency of the observed passband of the circuit in Fig. 2 and for the parameters x and K_{AB} in Fig. 4 can be derived in the same way as in [5]. The center frequency f_0 occurs where

$$X^eX^o=Z_0^2 \tag{6}$$

where Z_0 is the characteristic impedance of the uncoupled, terminating lines in Fig. 2(b). X^e and X^o are the reactances seen looking into one of the gratings under even- and odd-mode conditions, respectively. They can be computed using (2) and (3). Note that at the center frequency both Z^e and Z^o must be purely imaginary. The impedance inverter parameter K_{AB} can be found using [5, eq. (12)], with X^o being evaluated from (2) herein using the Θ^o fixed by (6) of this paper. The reactance slope parameter x is found through a similar procedure as in [5], and for microstrip gratings the result is

$$x=\frac{\Theta^o}{8}\left[\left(\frac{Z_0^e}{X^e}\right)^2\left(\frac{v^o}{v^e}\right)_{\text{ave}}h^e+\left(\frac{Z_0^o}{X^o}\right)^2h^0\right]\frac{\left(1-\left(\dfrac{K_{AB}}{Z_0}\right)^2\right)^2}{1+\left(\dfrac{K_{AB}}{Z_0}\right)^2} \tag{7}$$

where

$$h^{e,o}=\frac{(1+r^{e,o})(1-r^{e,o})\sin\Theta^{e,o}}{[(r^{e,o}+1)\cos\Theta^{e,o}-(r^{e,o}-1)]^2} \tag{8}$$

all evaluated at f_0 as fixed by (6).

It can be shown that the equivalent unloaded Q of the resonators in Fig. 4 is the same as that of a conventional resonator made from the lines comprising the gratings, i.e., given by

$$Q_u=\frac{\beta}{2\alpha} \tag{9}$$

where β is the (average) propagation constant of the lines and α is the (average) attenuation constant. This might be expected, but its proof is not obvious, especially in the case of coupled gratings. A derivation of this result can be found in [7].

Dispersion of the microstrip lines has been ignored in the foregoing discussion. Methods for incorporating corrections for dispersion were discussed in [5], but for typical microstrip lines the effects of dispersion on the shape and width of the passband are insignificant.

IV. On the Effects of Backward Coupling

The most serious, harmful effect of the so-called backward coupling (resulting from the difference in odd- and even-mode impedances of the line segments) is that it limits the maximum attenuation available from a pair of coupled gratings. If there were no differences in the odd- and even-mode impedances we would have $r^e=r^o$ and $Z_0^e=Z_0^o$. In that case it can be seen from (2a) that far from the center frequency of the coupled-grating passband Z^e and Z^o would approach the same value, implying that the absorptive attenuation of a pair of coupled gratings could theoretically be extremely large. This is more or less the case in dielectric waveguide gratings. Microstrip gratings, on the other hand, show less attenuation in the absorptive stopband because of the differing odd- and even-mode impedances.

If the separation between the gratings is increased, there is, of course, less coupling and the mode impedances differ less, as do the mode velocities. However, while the maximum backward coupling is rapidly diminished for loose couplings, the needed amount of forward coupling can still be obtained if the coupling length is long enough. Because of this, satisfactory performance can be obtained by a proper choice of the parameters of the coupled gratings. In filter design this may translate into compromising between stopband attenuation and maximum available filter passband width. Obviously, using a large separation between the gratings gives better stopband attenuation, but it can reduce the available filter bandwidth below what is required for some applications. For given gratings there is a unique grating spacing which will permit the maximum bandwidth [5]. However, the larger the impedance ratio r of the gratings, the larger that maximum possible bandwidth will be, the shorter the coupled gratings can be, and the lower the stopband attenuation will be (as a result of the presence of backward coupling).

We also found that the curved input strips of the coupled gratings are beneficial in reducing the unwanted backward coupling. Because the coupling between the input and output lines is continuously varied over a distance, they tend to match the odd- and even-mode impedances by acting as tapered line transformers and therefore reduce backward coupling while preserving any forward coupling. Because of this, higher stopband attenuations are achieved at frequencies at which the length of the curved strips is an appreciable fraction of a wavelength. This is readily apparent in the responses in Figs. 5 and 7. In the low-frequency limit attenuation is not much improved, however, because the curved strips become short compared to wavelength.

All these factors are therefore seen to be intricately interacting. It is difficult to make any generalized statements about the realizability of a design given some filter specifications or the optimum choice of structural parameters, as that depends on the required minimum stopband attenuation, passband width, filter prototype chosen, and microstrip substrate material used. The examples discussed

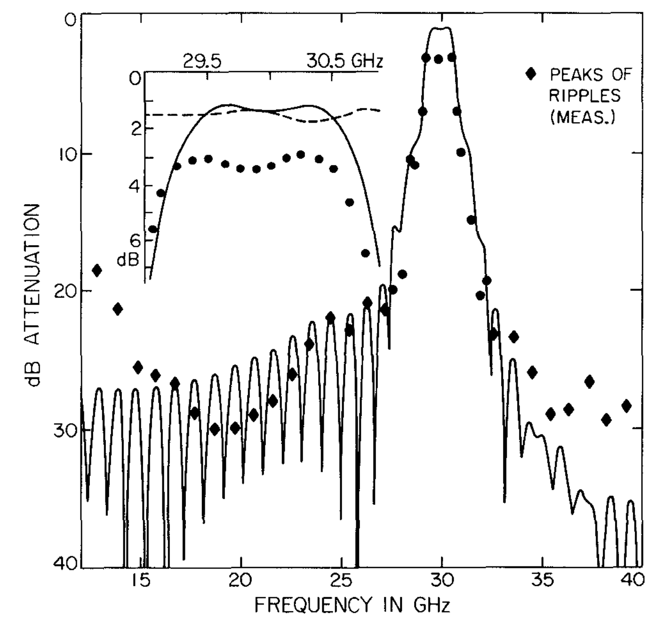

Fig. 5. A computed response for the 30-GHz, two-resonator filter. The dots indicate measured results, and the dashed line indicates the measured loss of the connectors and relatively lengthy input lines of the filter.

below should give some idea about practical designs. Stopband attenuations of about 20 dB or better (low-frequency limit) per pair of coupled gratings and filter bandwidths as wide as about 3.5 percent seem to be easily achievable.

V. Microstrip Coupled Grating Design Consideration

It is most convenient to start by fixing the widths W_0 and W_1 of Fig. 2(a). If the filter is to be operated with 50 Ω terminations W_0 should be chosen to give that impedance. W_1 becomes fixed by choosing the grating impedance ratio r. In our designs we have used impedance ratios 1.15 or 1.16 (this refers to the impedance ratio of the gratings if they were uncoupled) in filters with design bandwidths on the order of 3 percent. For narrower bandwidths, lower impedance ratios should be best (i.e., yield a design with a practical number of half-wavelengths in each resonator) while for wider bandwidths larger impedance ratios would be necessary. The spacing S in the parallel-coupled gratings has to be determined judiciously and may involve some computational trial and error and a tradeoff between stopband attenuation and filter bandwidth (if a wide passband width is desired), as explained above. The lengths l_0 and l_1 of Fig. 2(a) must be fixed so as to make the structure resonant at the center frequency of the filter. The resonant condition was given by (6). Another condition is needed to determine the two unknowns l_0 and l_1, and we have found it convenient to specify that l_0 and l_1 in Fig. 2(a) have to be equal in electrical length in the odd mode. As a practical matter, one could use $l_0 = l_1$ as they typically differ by only a minor amount.

Once the structural parameters of the coupled gratings have been fixed, the parameters of the equivalent circuit in Fig. 4 need to be computed. However, a practical circuit must use curved strips at the input to the coupled gratings while the theory was based on a "sharp-bend" model (see Fig. 2(b)). Our equations for K_{AB} and x in Fig. 4 can still be used to give initial estimates and are useful in studying their general dependence on the structural parameters of the coupled gratings. For precision filter design, however, we have found numerical simulation useful in determining final values of K_{AB} and x, and the exact position of the reference planes corresponding to A and B of Fig. 2(b) for coupled gratings with curved input strips.

For these computations the curved strips have been analyzed by dividing them into small segments which are treated as being parallel and uniformly coupled. Reference plane positions are signified by the property that at the center frequency a purely real load impedance Z_B at port B gets transformed into a purely real impedance Z_A at port A, and then $K_{AB} = \sqrt{Z_A Z_B}$. The location of the reference planes is quite accurately a quarter wavelength from the first notch of the coupled gratings, in accordance with the sharp-bend model of Fig. 2(b). Sometimes it may be advisable to fix the reference planes still a half wavelength further out on the curved strips because there may be nonnegligible coupling beyond the first location. The reactance slope x is found by leaving, say, port B open and evaluating $dX_{in}/d\omega$ at port A. Note that if the reference planes are fixed, say, a half wavelength from the first possible location, the x evaluated in the manner explained includes a contribution from these extra lines as well. These design computations involving gratings and curved lines were done very conveniently using the "Touchstone" commercial program.

VI. Design Examples and Experimental Results

We have designed, built, and tested several microstrip grating filters. A two-resonator and a four-resonator filter were designed for 10 GHz, and a two-resonator filter was tested at 30 GHz. The substrate material used was Duriod with a dielectric constant of 2.20. The substrate thickness was 30 mils for the four-resonator filter at 10 GHz and 11 mils for the 30 GHz filter, respectively. All of the filters had roughly 3 percent bandwidth. Plots in [2] and our Q calculation discussed in the introduction suggest that for the substrates and center frequencies that were used, conventional microstrip filters with half-wavelength resonators would not be practical for the given design bandwidths because of low resonator Q's caused by radiation. Metallization thickness was 1.3 mils for the 10 GHz and 0.7 mils for the 30 GHz designs, respectively. Losses and metallization thickness were ignored in most of the design calculations, as were fringing effects at the step discontinuities. For computing frequency responses of these filters we have used a program of our own as well as Touchstone. We found Touchstone to give improved accuracy because of a metal-thickness correction not included in our program. Agreement between measured and theoretical results was generally excellent. Details concerning the 30 GHz two-resonator filter and the 10 GHz four-resonator filter are given below.

A. Two-Resonator Filter at 30 GHz

The filter was designed using methods in [5] and in the preceding sections herein to have a 0.5-dB Chebyshev ripple response with 2.5 percent bandwidth. As we have consistently found, the actual bandwidth of the design came out to be somewhat larger than the initially specified bandwidth (3.3 percent actual bandwidth in this case). (Details of the design may be found in [7].) The resulting strip pattern is approximately as shown in Fig. 3. The impedance ratio used in the coupled gratings is $1.15 = 57.5/50.0$ (refers to the impedances of the lines as uncoupled). A few different values of spacing between the coupled gratings were tried computationally, and a spacing equal to the substrate thickness seemed to offer a reasonable compromise for our purposes. The coupled gratings had 30 Z_1 segments. (i.e., notches). The G_1 gratings in Fig. 3 had four notches as shown (though the notches were much more shallow), and n in the figure is two. In order to simulate infinitely long gratings the portion covering the six last segments (off to the right in Fig. 3, not shown) had distributed loads in the form of microwave absorbing material added next to the lines. A 50 Ω chip resistor was attached to the ends. The distributed loads were modeled in computations by adjusting the loss tangent of the substrate. The curved strips in Fig. 3 have a radius of curvature of 0.63 in, and 20 segments, each 0.010 in long, were used in the computations to model them. The length of this filter was 5.2 in excluding connectors and input lines.

It was found that the filter might be slightly mistuned if it is designed ignoring metallization thickness effects. These effects were easily compensated by making the two half wavelengths of Z_0 between the gratings slightly longer to correct for the center frequency, while the ripple size was corrected by a slight increase in the impedance ratio of the input and output coupling gratings (which has the effect of increasing the external Q of the resonators). Theoretical responses (including metal thickness effects and losses) for the final filter design are shown by the solid lines in Fig. 5.

A corresponding experimental filter was built and tested. The filter utilized K-connector "sparkplug" coax-to-microstrip launchers, which are rated for operation up to 40 GHz. However, for measurements in the 18–40 GHz band waveguide hardware was used together with coax-to-waveguide adapters in order to make the transition to the coaxial connectors of the filter. Measured results (including the launchers and some microstrip line) are shown by dots in Fig. 5 while the diamond points in the figure indicate the measured peaks of the stopband ripples. A test was conducted on a reference line to determine the loss due to the transition and microstrip input lines, and their measured loss is shown by the dashed line in the inset of Fig. 5. After the loss of the reference line is subtracted from the measured passband loss, the loss of the filter alone is found to be only slightly higher than predicted theoretically (less than 0.3 dB difference), showing that these filters, indeed, radiate very little and can take advantage of the full potential Q of microstrip lines. Note also that the filter has a larger ripple than theoretically predicted, which tends to increase the losses of the experimental filter.

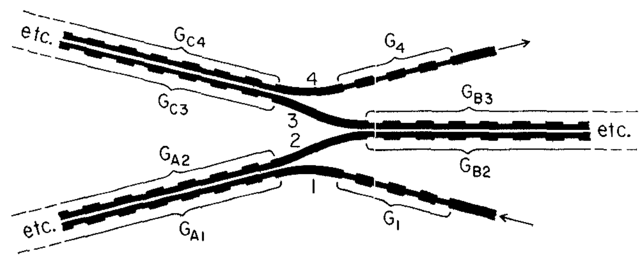

Fig. 6. A partial drawing of the 10-GHz, four-resonator filter. The notch depth of the gratings has been exaggerated.

B. Four-Resonator Filter at 10 GHz

The configuration of this filter is somewhat different from the four-resonator filter proposed in dielectric waveguide form [5]. A partial drawing of the filter is shown in Fig. 6. In this filter the coupling of the first and last resonators to the terminations is via single gratings, as in the four-resonator filter of [5], but all interresonator couplings, including the coupling between the second and the third resonator, are arranged entirely through pairs of parallel-coupled gratings. (The four-resonator filter in [5] used a cascaded grating in its center and only two pairs of parallel-coupled gratings.) Thus, the filter in Fig. 6 has three pairs of coupled gratings. Since the attenuation per pair is absorptive over most of the stopband, their dB attenuations add, giving the filter very potent stopband characteristics. The design procedure for this filter configuration is slightly different from that given in [5] and can be found in full detail in [7]. This approach could also be used for design of a three-resonator filter having two pairs of coupled gratings.

The filter shown in Fig. 6 was designed from a 0.2-dB ripple prototype for a 2.9 percent bandwidth (which enlarged to 3.4 percent in the actual design). The impedance ratio used in the coupled gratings is 1.16, and a spacing of 1.2 times the substrate thickness was used between them. Again, 30 notches were used in the parallel-coupled gratings. Computed responses are shown by the solid lines in Fig. 7. The effects of metallization thickness (and losses) have been included in these computations while the initial design ignored these effects. It can be seen from the inset of Fig. 7 that the computed response predicts that the filter is slightly mistuned because of the effects of the metallization thickness (in this case no corrections for finite metal thickness were made in the design).

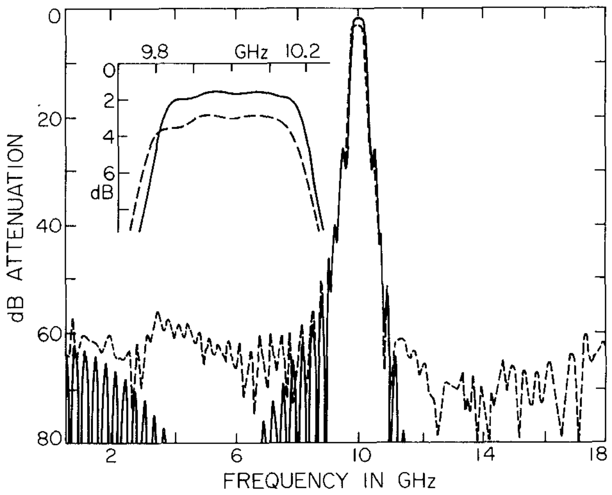

Fig. 7. Computer (—) and measured (---) responses for the 10-GHz, four-resonator filter shown in Fig. 6. The measured data include the loss of extra lengths of line.

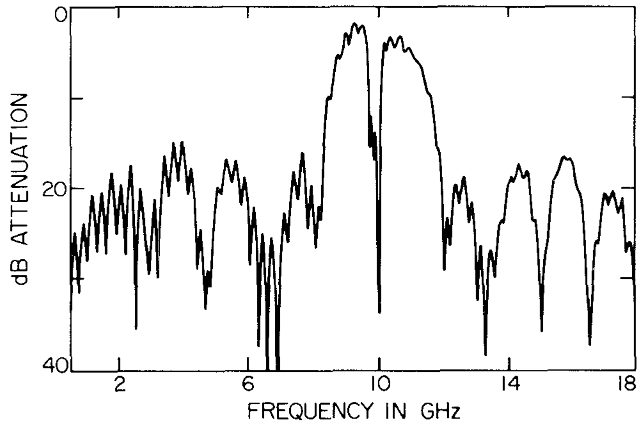

Fig. 8. Measured dB return loss at the input of the filter in Figs. 6 and 7.

Corresponding measured results are shown by the dashed lines in Fig. 7. It should be noticed that the measured result again includes the losses of input lines and connectors. Their contribution is about 1.0 dB, so the measured response of the filter alone is in excellent agreement with the computed one. In the stopband it was found that the attenuation was limited primarily by cross talk between different parts of the filter rather than by reaching the limit indicated as possible by the theoretical calculations (see Fig. 7). To get the response shown in Fig. 7 the coax–microstrip transitions were shielded with conductive covers and absorbing material was used between gratings G_1 and G_{B2} in Fig. 6 (as well as between gratings G_4 and G_{B3}) and between gratings G_{A2} and G_{C3} in order to isolate these gratings from each other. On the other hand, the absorptive shieldings were kept sufficiently far away from the lines so as not to increase passband loss. The additional loss due to them was measured to be only about 0.2 dB over that shown in the inset of Fig. 7. However, even without any shielding the minimum stopband attenuation was measured to be better than 50 dB at all frequencies in the band 0.5 to 18 GHz, except at 11.3 GHz, where there was a glitch at which the attenuation reduced to 40 dB. Thus, even without any shielding inserts the stopband performance of these filters is relatively strong. The next passband for this filter should occur about 30 GHz. Additional measurements were made up as far as 26.5 GHz with the absorptive shielding in place. The measured attenuation was in excess of 60 dB except for dips in the vicinity of 19.4 and 20.4 GHz, where the attenuation dropped to about 56 dB. We believe that the broad, strong stopbands of this type of filter should make them particularly attractive for some millimeter-wave applications. (For example, *E*-plane filters have a second passband centered less than an octave above the first passband.)

Some effort was made to compare the rate of cutoff of grating filters with that of comparable filters of other types. *E*-plane filters are the main millimeter-wave filters available for narrow-band bandpass applications. We attempted to draw conclusions from a comparison of the 2.8% bandwidth, five-resonator *E*-plane filter response in [9] with our 3.4% bandwidth four-resonator grating filters. Due to the disparity between the filters precise conclusions were impossible to make; however, it appears that for frequencies at the edges of the passband to where the first sidelobes appear in Fig. 7 (at about the 23-dB level) the performance of both types of filters is probably very similar. However, beyond the first side lobes the rate of cutoff for comparable grating filters is slower.

Measurements were made of the return loss from 0.5 to 18 GHz, and the results are shown in Fig. 8. As is to be expected, the return loss is small on both sides of the passband where the gratings are still reflecting and the filter is acting as a reflection-type filter. The effect of the mistuning due to ignoring metallization thickness in the design is evident in the passband region. The form and details of the computed response (not shown) are very similar to the measured data.

VII. Conclusions

The coupled-grating filter structure has been found to be adaptable for use with microstrip. Narrow-band bandpass filters can be obtained with very little radiation when using substrates at frequencies for which half-wavelength resonators would radiate strongly. Since there is little radiation and since relatively thick substrates can be used, comparatively low passband loss is achieved. On the other hand, the coupled gratings give these filters wide and relatively strong stopbands, with the attenuation being absorptive at most frequencies. It is believed that comparable performance (narrow bandwidths with low loss and wide and strong stopbands) would be difficult to achieve with other microstrip filter configurations. The main disadvantage of these filters is that the coupled gratings have to be around 6 to 8 wavelengths long, so they tend to be relatively large. However, at millimeter wavelengths, where their low-radiation characteristics are of most value, their overall size should be reasonable for many applications, such as hybrid integrated circuits. Another limitation is that the passband cannot be very wide, and the wider it is, the less the absorptive attenuation in the stopband because of the backward coupling between the microstrips, as was explained. This should not be a serious drawback since low radiation is mostly needed for narrow-band designs where the high passband loss of conventional microstrip filters is not acceptable. We found bandwidths up to about 3.5 percent to be easily achievable with the absorptive stopband attenuation of a pair of coupled gratings still being about 20 dB at low frequencies (and significantly more at high frequencies due to a matching effect, as was explained). Also, multiresonator designs with more than one pair of coupled gratings can be used to enhance the stopband attenuation.

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