United States Patent

[11] 3,634,788

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[45] Patented Jan. 11, 1972

[73] Assignee International Standard Electric Corporation
New York, N.Y.

[54] WAVEGUIDE FILTER
4 Claims, 44 Drawing Figs.

[52] U.S. Cl. 333/73, 333/83, 333/81
[51] Int. Cl. H03h 13/00, H03h 7/10
[50] Field of Search ..... 333/73, 73 C, 73 W, 10, 7, 9

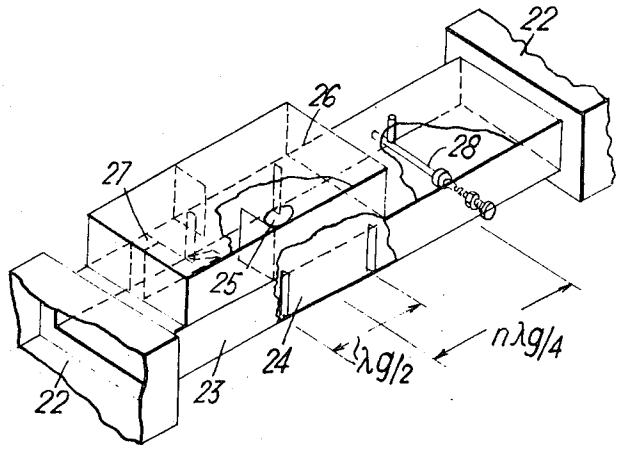
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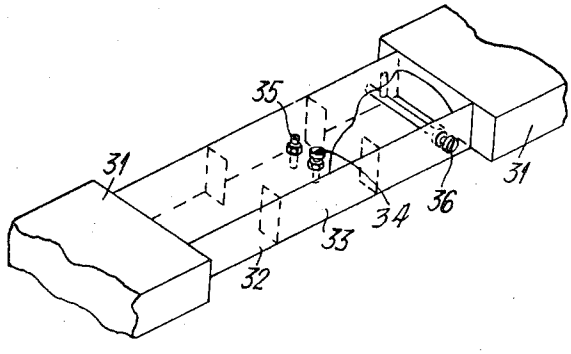
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Primary Examiner—Herman Karl Saalbach
Assistant Examiner—C. Baraff
Attorneys—C. Cornell Remsen, Jr., Rayson P. Morris, Percy P. Lantzy, Philip M. Bolton and Isidore Togut

ABSTRACT: This invention relates to waveguide filters wherein waveguide sections functioning in the evanescent mode for frequencies in the desired passband are utilized to couple conventional cavity filter sections together. The evanescent sections operate in their normal modes at frequencies higher than the passband frequencies. Suppression devices are then coupled to one or more of the evanescent sections (instead of to the cavity filter sections) to suppress the parasitic harmonic waves while having negligible effect on the evanescent mode passband frequencies.





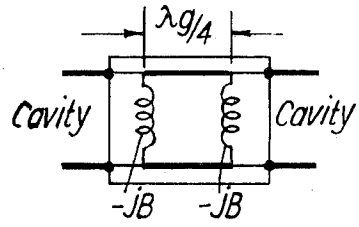


Fig. 1(a).

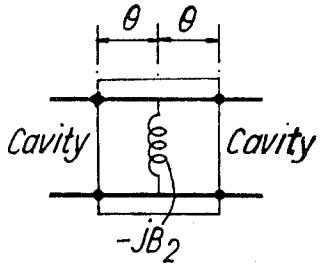


Fig. 1(b).

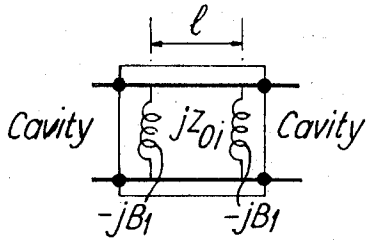


Fig. 1(c).

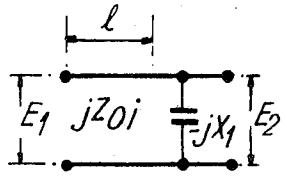


Fig. 1(d).

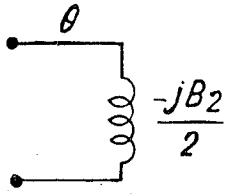


Fig. 2(a).

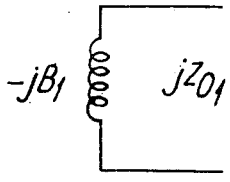


Fig. 2(b).

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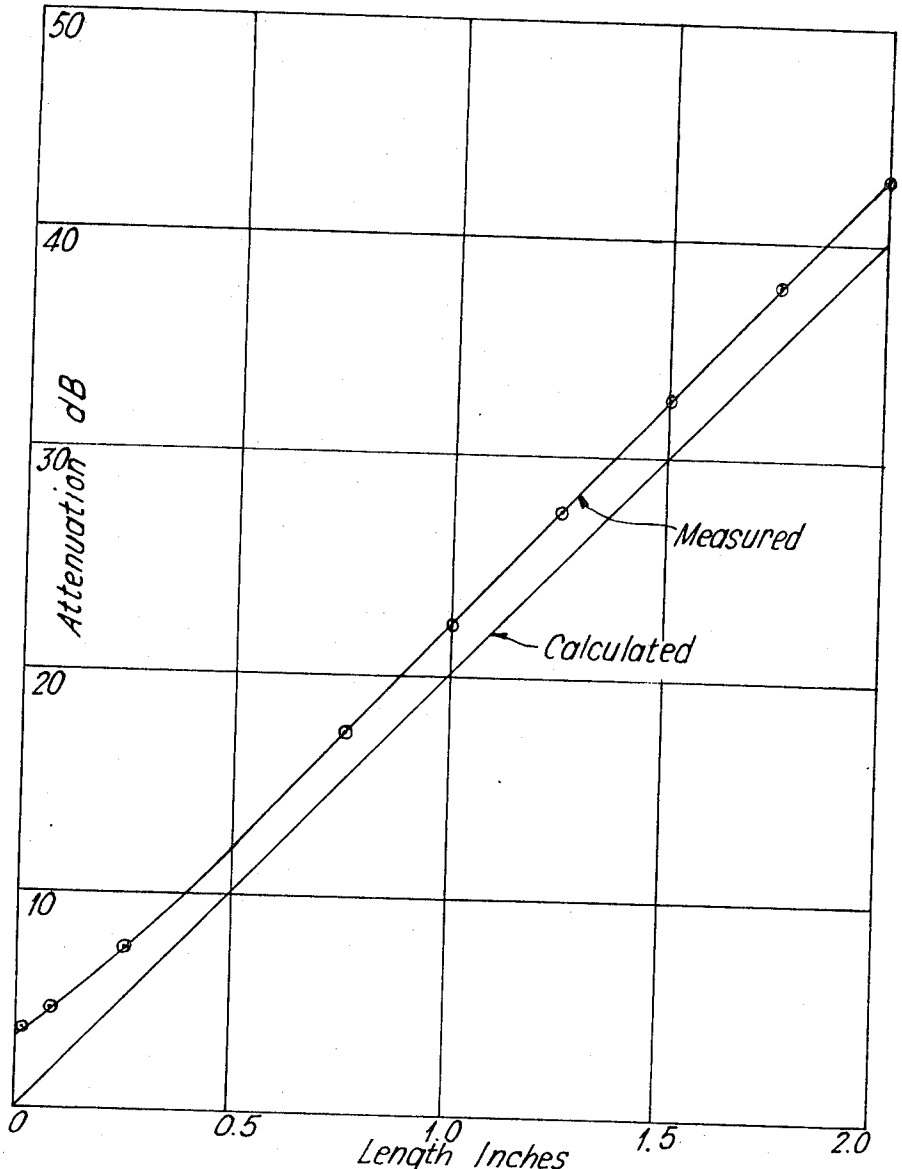
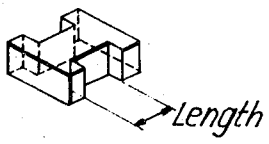


Fig.3.

Fig.4.



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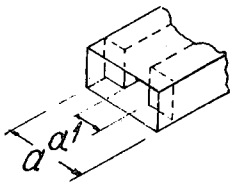


Fig. 5(a).

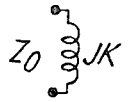


Fig. 5(b).

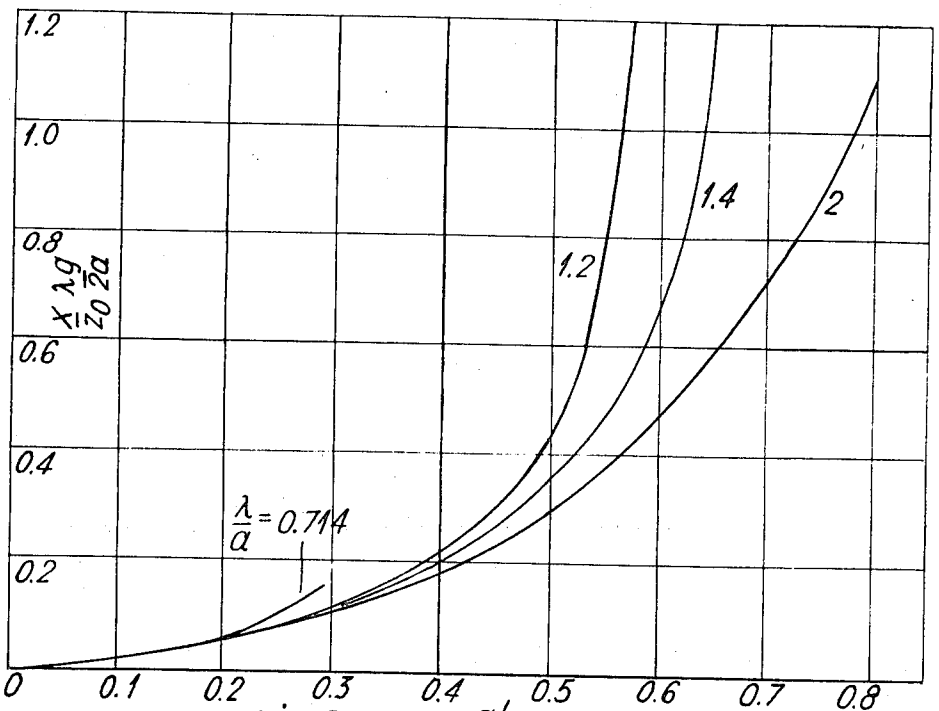


Fig. 6.

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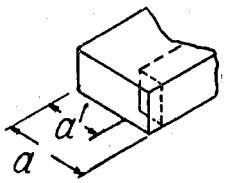


Fig. 7(a).

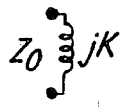


Fig. 7(b).

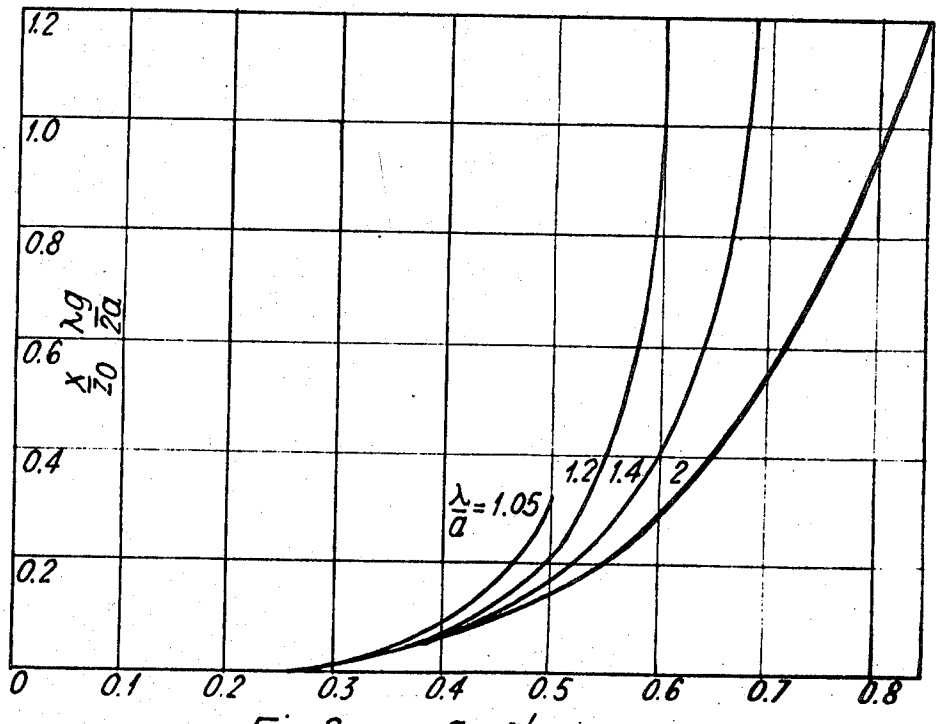


Fig. 8.

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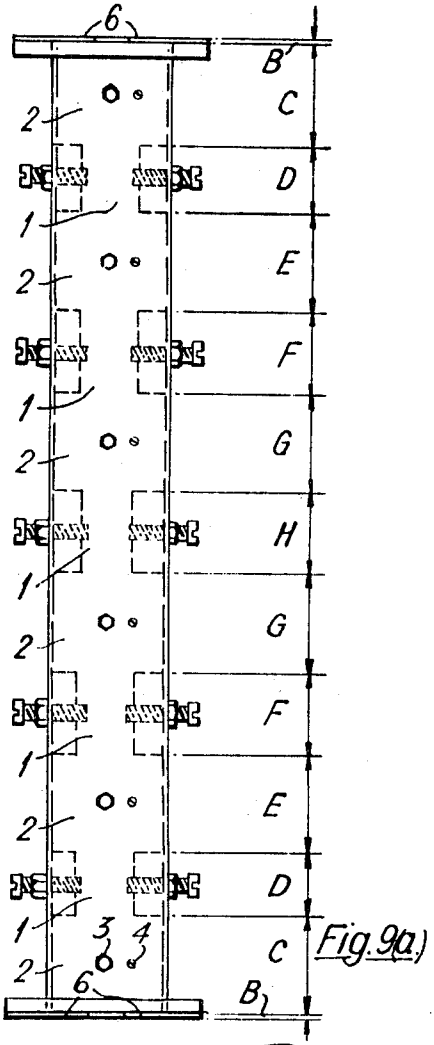


Fig. 9(a)

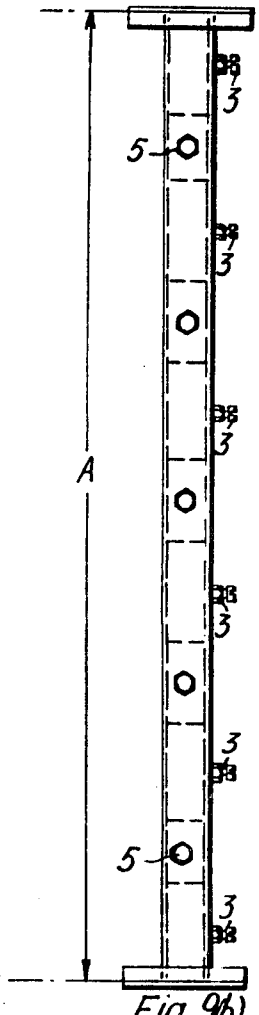


Fig. 9(b)

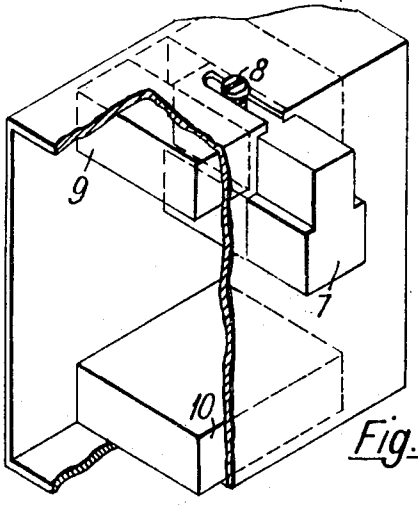


Fig. 10

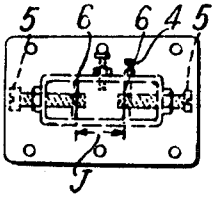


Fig. 9(c)

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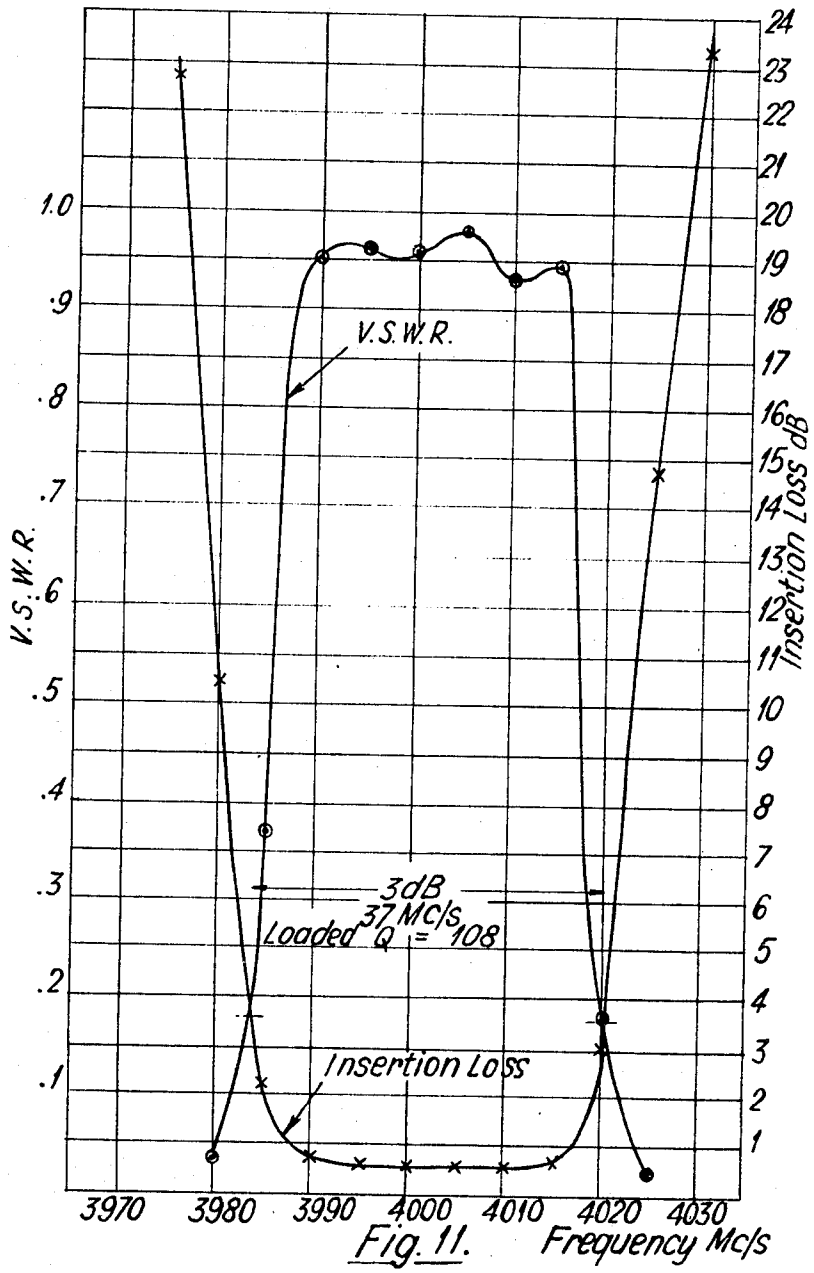


Fig. 11.

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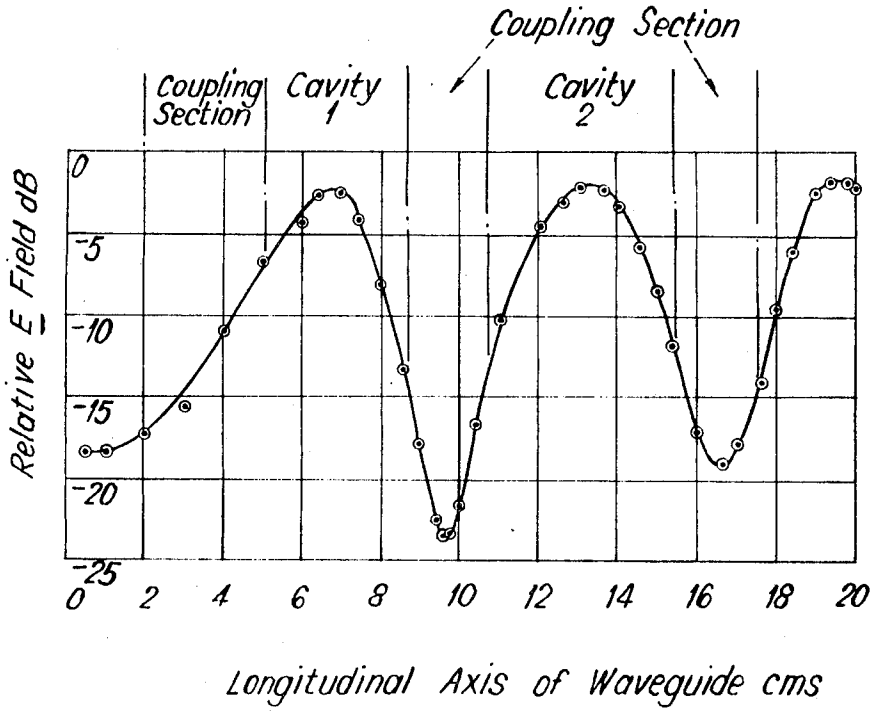


Fig. 12.

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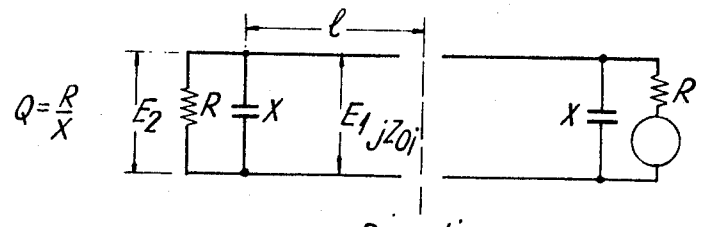


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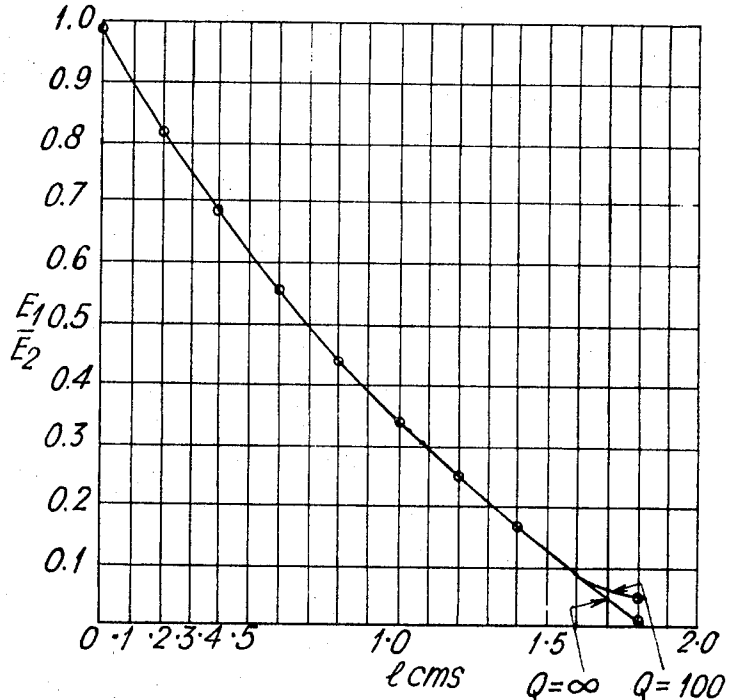


Fig. 14.

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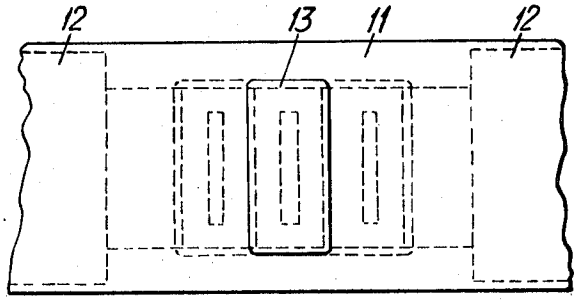


Fig. 15(a).

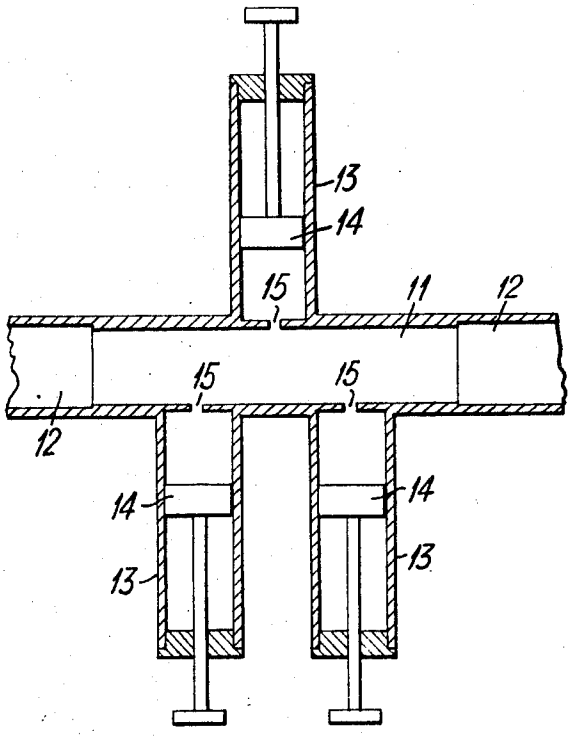


Fig. 15(b).

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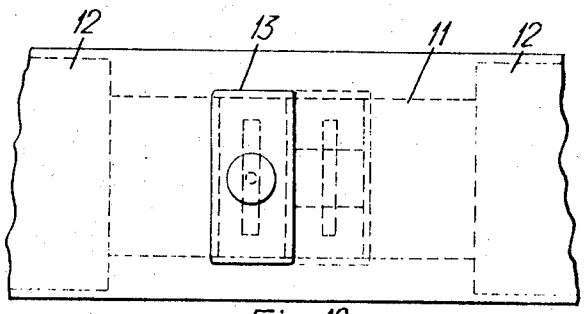


Fig. 16(a).

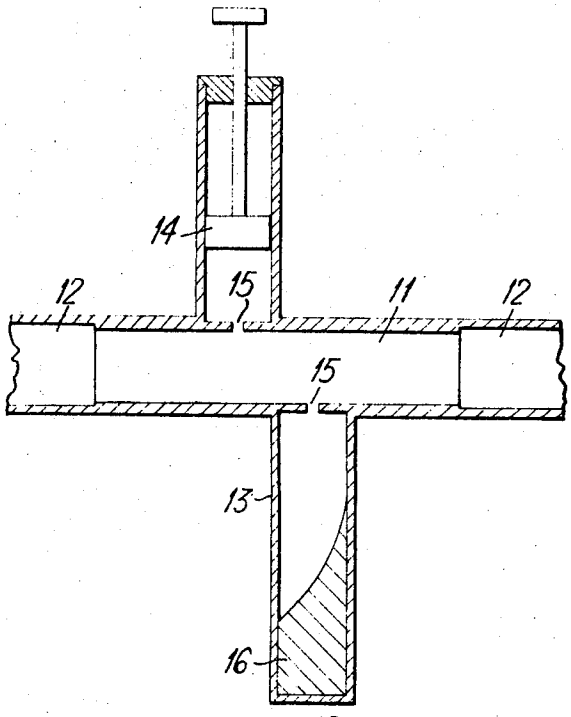


Fig. 16(b).

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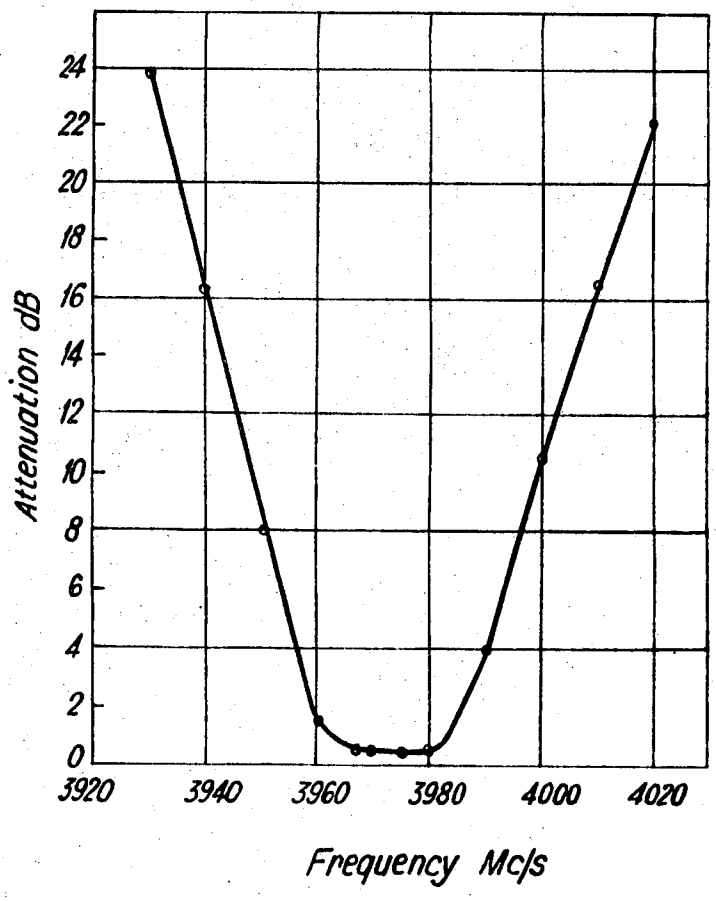


Fig. 17.

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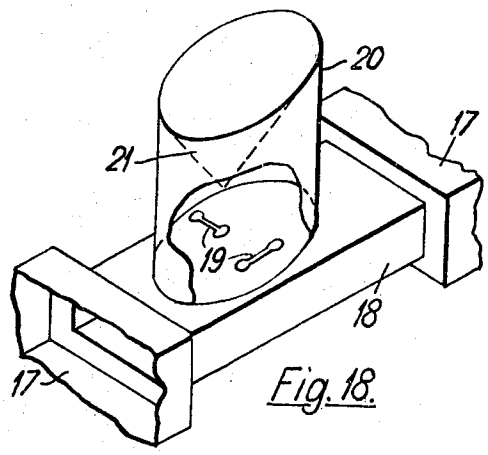


Fig. 18.

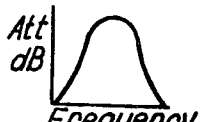


Fig. 19.

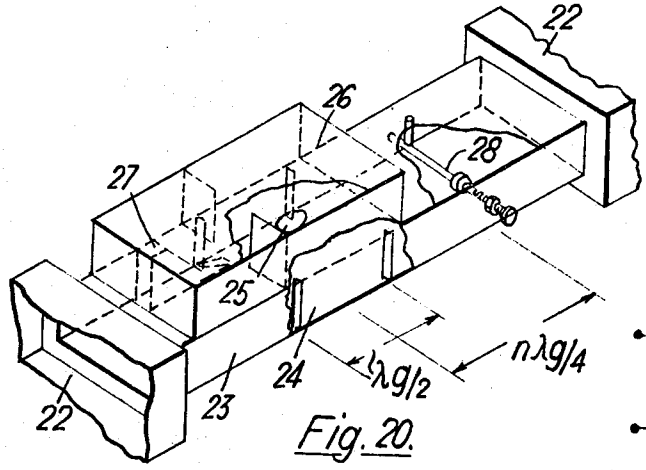


Fig. 20.

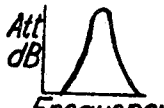


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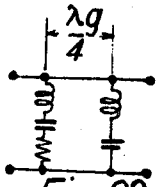


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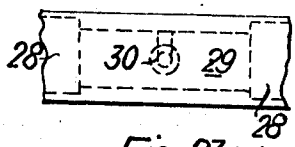


Fig. 23(a).

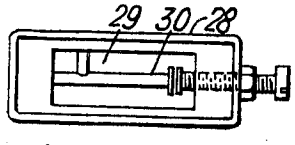


Fig. 23(b).

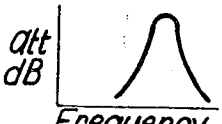


Fig. 24.

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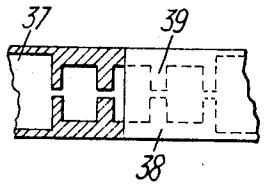


Fig. 27(b).

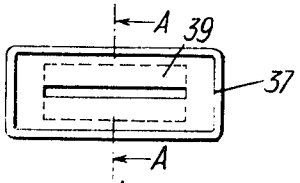


Fig. 27(a).



Fig. 28.

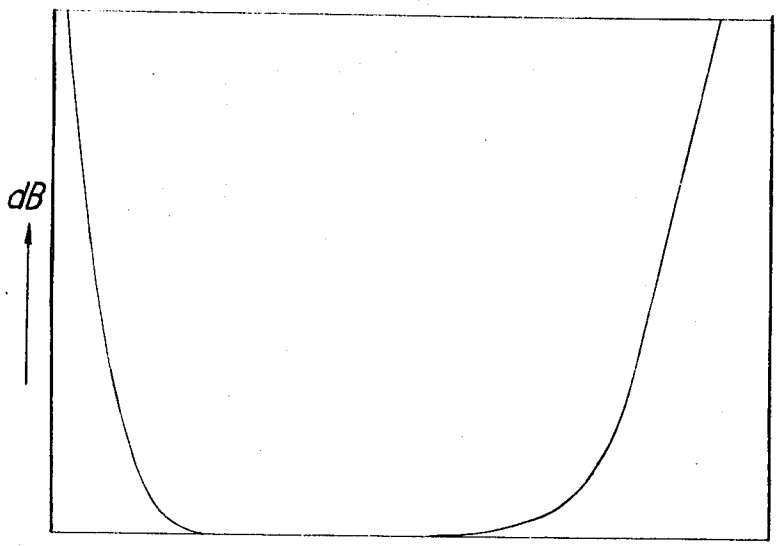


Fig. 29.

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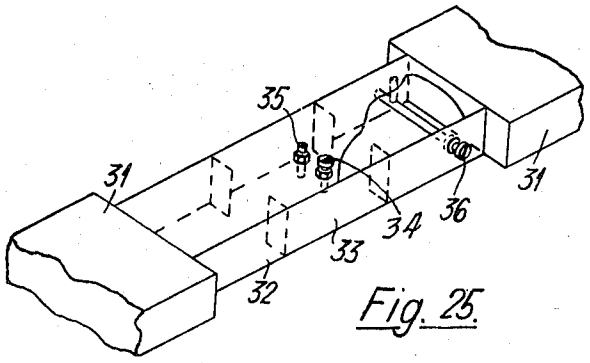


Fig. 25.

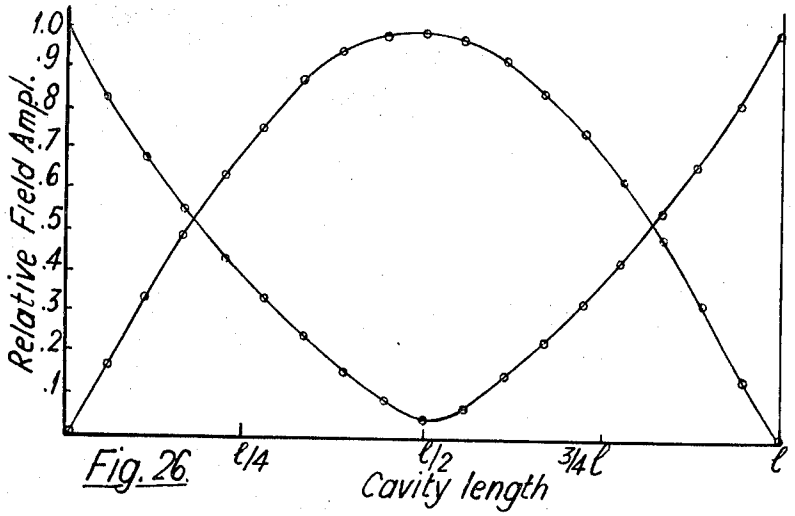


Fig. 26.

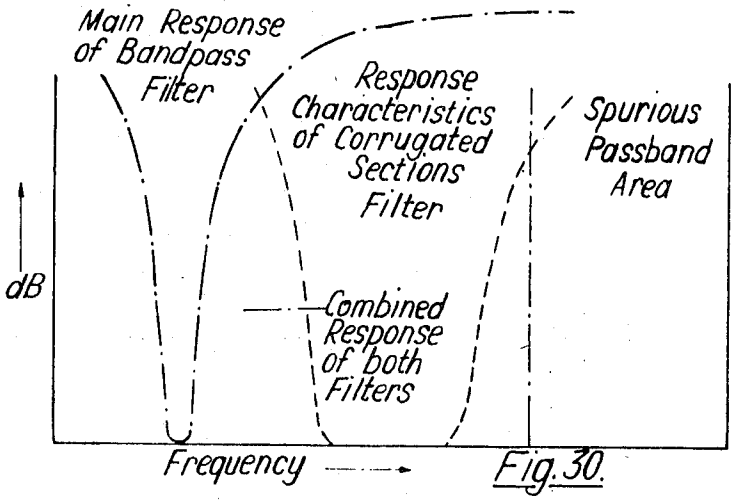


Fig. 30.

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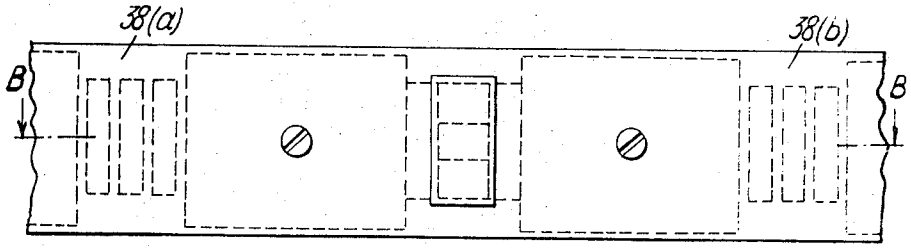


Fig. 31(a).

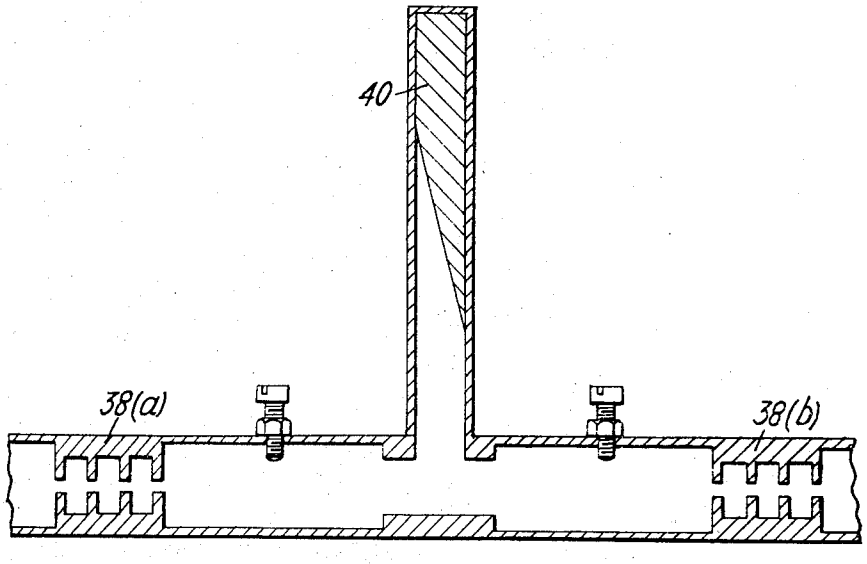


Fig. 31(b).

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WAVEGUIDE FILTER

BACKGROUND OF THE INVENTION

This invention relates to waveguide band-pass filters.

Conventional direct-coupled and quarter-wave-coupled filters each have advantages and disadvantages. The advantages of the quarter-wave-coupled filter lies in its small coupling susceptances and consequently, the ease with which a paper design can be realized with practical mechanical tolerances. Initially, this type of filter was preferred because the cavities could be made separately and tuned before assembly. However, now that filters are always made and tuned as a complete unit this advantage is of little value and general, if not complete, preference has passed to the direct-coupled filter. The direct-coupled filter eliminates the frequency-sensitive quarter-wave couplings and, therefore, is applicable to wider bandwidth designs; it is also shorter (about 75 percent of the length of the quarter-wave filter). This is achieved by substituting one large susceptance for the two smaller susceptances and the quarter-wave coupling between cavities. This makes the direct-coupled filter potentially cheaper although this is offset by the much stricter tolerances, or additional adjustments, that are necessary. The difficulties arise from the highly critical nature of large susceptances which, in the example of a symmetrical diaphragm, for instance, varies as $\cot^2 (\pi d_1/2a)(d_1 \to 0$ for large susceptances). Use has also been made of multipost arrangements but this alternative leads to manufacturing difficulties and the thin posts also increase loss.

SUMMARY OF THE INVENTION

Therefore, the main object of this invention is to provide an improved waveguide filter wherein the higher order parasitic frequencies are suppressed while having negligible effects on the desired passband signals.

According to the invention there is provided a waveguide band-pass filter having main resonator cavities coupled together by waveguide sections which are evanescent at the passband frequency of the filter.

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. 1(*a*) is the equivalent circuit of two cavities with a conventional quarter-wave coupling,

FIG. 1(*b*) is the equivalent circuit of two cavities with a conventional direct coupling,

FIG. 1(*c*) is the equivalent circuit of two cavities with an evanescent mode coupling,

FIG. 1(*d*) is the voltage transfer equivalent of an evanescent section,

FIG. 2(*a*) is the bisected circuit of FIG. 1(*a*).

FIG. 2(*b*) is the bisected circuit of FIG. 1(*c*).

FIG. 3 shows a typical attenuation/length characteristic for an evanescent mode coupling measured at 4,000 mc./sec.,

FIG. 4 shows the evanescent mode coupling giving the characteristic of FIG. 3,

FIGS. 5(*a*) and 5(*b*) show a symmetrical evanescent mode coupling its equivalent circuit, respectively,

FIG. 6 shows the reactance at a junction of a waveguide terminated in the symmetrical evanescent section of FIG. 5(*a*).

FIG. 7(*a*) and 7(*b*) show an asymmetrical evanescent mode coupling and its equivalent circuit, respectively,

FIG. 8 shows the reactance at a junction of a waveguide terminated in the symmetrical evanescent section of FIG. 7(*a*),

FIGS. 9(*a*), 9(*b*) and 9(*c*) are plan, side and end views, respectively, of a six-section resonant cavity waveguide band-pass filter with evanescent mode coupling sections,

FIG. 10 is a perspective view, partially cut away, of a modified form of an evanescent mode coupling section,

FIG. 11 shows the curves insertion loss and V.S.W.R. vs. frequency for the filter of FIG. 9,

FIG. 12 shows the measured standing wave pattern in a three-section filter with evanescent mode coupling sections throughout,

FIG. 13 shows the bisected equivalent circuit of a symmetrical network of imaginary characteristic impedance terminated for full energy transfer,

FIG. 14 shows the calculated standing wave pattern for the symmetrical half-section evanescent coupling between cavities of FIG. 13,

FIGS. 15(*a*) and 15(*b*) are plan and sectioned side views, respectively, of an evanescent coupling section with a harmonic rejection filter,

FIGS. 16(*a*) and 16(*b*) are plan and sectioned side views respectively of an evanescent coupling section with a constant resistance rejection filter,

FIG. 17 shows the transmission response of the filter of FIG. 16,

FIG. 18 is a perspective partially cutaway view of an evanescent coupling section with a resonant slot hybrid junction rejection filter,

FIG. 19 shows the attenuation characteristic of the filter of FIG. 18,

FIG. 20 is a perspective partially cutaway view of an evanescent coupling section with a form of constant resistance rejection filter,

FIG. 21 shows the attenuation characteristic of the filter of FIG. 20,

FIG. 22 shows the equivalent circuit of the filter of FIG. 20,

FIGS. 23(*a*) and 23(*b*) are side and end views, respectively, of an evanescent coupling section with a series resonant rejection filter,

FIG. 24 shows the attenuation characteristic of the filter of FIG. 23,

FIG. 25 is a perspective partially cutaway view of an evanescent coupling section with a form absorption filter,

FIG. 26 shows the standing wave pattern in the coupling section of FIG. 25,

FIG. 27(*a*) is an end view, and FIG. 27(*b*) is a half-sectioned side view along the line A—A of FIG. 27(*a*), of an evanescent coupling section with a low-pass rejection filter,

FIG. 28 shows the attenuation characteristic of the filter of FIG. 27,

FIG. 29 shows a typical frequency characteristic of a waveguide low-pass filter,

FIG. 30 shows the wide-band rejection characteristics of a waveguide low-pass filter, and

FIG. 31(*a*) is a plan view, and FIG. 31(*b*) a section on the line B—B of FIG. 31(*b*), of cascaded low-pass filter evanescent coupling sections with a dissipative section in between.

DESCRIPTION OF THE PREFERRED EMBODIMENTS

Waveguide at frequencies below cutoff exhibits characteristics common to all nondissipative filter networks in their stop-band region. The characteristic impedance, which is real in the passband becomes imaginary in the stop-band. The propagation constant, which is imaginary in the passband becomes real in the stop-band. The transmission line analogue then has the properties shown in FIG. 1(*c*), FIGS. 1(*a*) and 1(*b*) represent the corresponding circuits for quarter-wave and direct-coupling, respectively. The conditions for virtual identity between the two circuits of FIGS. 1(*a*) and 1(*b*) have been established for example, in "An Improved Design Procedure of the Multisection Generalized Microwave Filter," Levy R., I.E.E. Monograph Vol. 232R, Apr. 1957, in which it is shown that if the insertion loss and phase shift of the two networks are the same, the circuit of FIG. 1(*a*) may be replaced by that in FIG. 1(*b*). In a similar way the network of FIG. 1(*c*) may be replaced by that in FIG. 1(*b*). The input reactance of the network of FIG. 1(*b*) is equivalent to a certain line length in the network of FIG. 1(*c*) and thus, the cavity length will be generally less when evanescent couplings are used. The insertion loss of the two networks may then be made identical by appropriate choice of the parameters in the evanescent section.

The conditions for equivalence between the two networks of FIGS. 1(*b*) and 1(*c*) can be readily derived from the standard transmission line equations. As the networks are symmetrical they may be bisected (FIGS. 2(*a*) and 2(*b*) nd the open- and short-circuit admittances derived. If the respective open circuit and short-circuit admittances can be equated and solved for physically realizable values then the two networks may be made identical at a given frequency. The short-circuit admittances for the networks of FIGS. 2(*a*) and 2(*b*) are, respectively.

$$Y_{sc1}=-jYo\cot\theta \quad (1)$$

$$Y_{sc2}=-jB_1-jYoi\coth(\nu l/2) \quad (2)$$

The open-circuit admittances for the same networks are:

$$Y_{oc1}=-jYo\cot(\nu+\phi) \quad (3)$$

where $\phi=\cot^1(B_2/2Yo)$

$$Y_{oc2}=-jB_1-jYoi\tanh(\nu l/2) \quad (4)$$

Equating the short-circuit admittances.

$$Yo\cot\theta=B_1+Yoi\coth(\nu l/2) \quad (5)$$

Equating the open-circuit admittances.

$$Yo\cot(\theta+\phi)=B_1+Yoi\tanh(\nu l/2) \quad (6)$$

Expanding the left hand side (6) and substituting for cot θ (from (5)) we have, solving for cot θ,

$$\cot\phi=\frac{1+\frac{B_1^2}{Yo^2}+\frac{Yoi^2}{Yo_2}+\frac{B_1Yoi}{Yo^2}\tanh\frac{\gamma l}{2}+\frac{B_1Yoi}{Yo^2}\coth\frac{\gamma l}{2}}{\frac{Yoi}{Yo}\left(\coth\frac{\gamma l}{2}-\tanh\frac{\gamma l}{2}\right)} \quad (7)$$

The extreme values that cot ϕ can have are when $\gamma l \rightarrow \infty$ and $\gamma l \rightarrow 0$.

In the first case both $\coth\frac{\gamma l}{2}$ and $\tanh\frac{\gamma l}{2}$ tend to unity. Thus, cot $\phi \rightarrow \infty$ as the line length tends to infinity which is the expected result. In the second case as $\gamma l \rightarrow 0$ the expression reduces to:

$$\cot\phi=(B_1/Yo) \text{ or } (B_2/2=B_1$$

which is obviously true. Equation (5) states simply the second requirement for equivalence i.e., the relationship between the line length in FIG. 2(*a*) and the combined lumped susceptance including the junction susceptance and the (inductive) characteristic admittance of the line.

The notion of energy propagating freely through a nondissipative line (or guide) of imaginary characteristic impedance, with the propagation constant real and assuming large values, may conflict with the intuitive notions held by many. This concept amounts to the belief that the ratio of input voltage, E_1, to output voltage, E_2, is given to good accuracy by the equation:

$$(E_1/E_2)=e\nu^l$$

where

γ = propagation constant

l = length

However, this is only true if the network is terminated in its characteristic impedance (imaginary). With other terminations the results can be very different. For instance, if the termination is capacitive (FIG. 1*d*) the ratio E_1/E_2, is given by

$$E_1/E_2=\cosh\nu l-(Zo_1/x_1)\sinh\gamma l \quad (8)$$

For large values of γl, cosh γl and sinh γl are nearly equal. Thus, for example, if $Zo_1\approx x_1$, E_1/E_2 is quite small and may be zero. The circuit then behaves as a series resonant circuit with a large resonant rise in voltage across the "inductance" and capacitance. Clearly, a wide range of values including $E_1/E_2=1$ is obtainable. A general engineering conception of the behavior is simply to say that the reactive insertion loss of evanescent guide may be "tuned out" by a suitable termination. Such evanescent waveguide sections are generally described on my copending application, Ser. No. 643,279. It roughly corresponds to neutralizing the effect of a shunt susceptance by adding one of equal value (but opposite sign) in shunt with it.

The design of direct-coupled filters is described in "Direct Coupled Filters," Cohn S. B., Proc. I.R.E., Feb. 1957. The

coupling parameter between resonant cavities is a common shunt susceptance. However, the design of a filter using evanescent couplings is more conveniently handled in terms of the insertion loss of the coupling section. The insertion loss of a susceptance shunted across a matched line is given by:

$$L_{db}=10\log_{10}\frac{(2+b^2)^2}{4(1+b^2)} \quad (9)$$

which for large values of b is, to good accuracy,

$$10\log_{10}\frac{3+b^2}{4}$$

The insertion loss of a section of evanescent guide interposed between matched guides will be a function of the guide length and junction susceptance. The guide length will normally be the principal factor, the insertion loss being:

$$L=e^{\gamma l}$$

or

$$Ldb=8.686\gamma l$$

where

$$\gamma=\frac{2\pi}{\lambda c}\sqrt{1-\left(\frac{\lambda^c}{\lambda^o}\right)^2} \quad (10)$$

and λc = cutoff wavelength λo = free space wavelength.

The junction susceptance is generally quite large but the insertion loss from this effect is less than might be expected. The reason for this is that the field from the junction spreads into the evanescent section causing the initial rate of decay to be less than $e^{\gamma l}$, FIG. 3 shows the insertion loss, as a function of length, of a typical section shown in FIG. 4 of 1.000×0.667-inch waveguide 41 coupling 2.000×0.667-inch waveguide 42. Except when the evanescent section is very short (and ultimately degenerates into a thin iris) the curve is a straight line. Thus, only two properly chosen measurements are essential in order to establish a design curve. Comparing the calculated (eq. (10)) and measured curves it will be seen that they are approximately 2 db. apart and parallel within about 1 db. over the range. Therefore, eq. (10) will give a very accurate guide to the effect that minor variations in λ_c will have on the measured attenuation curve. It is of interest that the difference between the two curves is roughly one-half the insertion loss of an iris of the same dimensions as the cutoff guide. This is a useful rule of thumb correction, if it is desired to use the theoretical curve for ratios a'/a (a'/a is defined in FIGS. 5 to 8).

The practical design of a filter is carried out in the following way. The values of the prototype sections are calculated using the equations in FIG. 2 of Cohn's paper. The susceptance values are then determined from FIG. 5 of the same paper. The insertion loss is then calculated from eq. (9) of this specification. A width dimension, which is well beyond cutoff at the passband frequency is chosen, and the input reactance of the section is then determined (FIG. 6 or 8). The "half-length" of the cavity is then found from:

$$\theta_2=(\pi/2)-\phi \quad (11)$$

Where $\phi=\tan(-1_x/Zo)$ (see FIG. 6 or 8 for value of (x/Zo). If the cavity is physically symmetrical about the electrical center i.e., the input reactance at each end is identical) the total length will be twice the value found in eq. (11); otherwise the second electrical length is calculated. If the total cavity length is unreasonably short as a consequence of the end loading, a second trial using a smaller guide width for the evanescent section will be necessary. The length of the section which will yield the desired insertion loss is then found from a curve such as that shown in FIG. 3.

Where the precise value of the filter bandwidth is not very important an approximate design technique using eq. (10) with an approximate correction for the discontinuity, can be employed. The desired characteristic maximally flat or Tchebycheff—will be obtained even if the theoretical attenua-

tion curve is in error. This follows because the correct ratio between the insertion loss values will be realized. The overall error in bandwidth can be gauged roughly by noting that an error of 0.5 db. in all of the intercavity coupling sections will produce an error of about 5 percent in the bandwidth of a typical filter.

So far only the intercavity couplings have been discussed as it is in these sections that evanescent couplings have their chief application. Susceptances at the end of the filter are usually of a comparatively low value and may, conveniently, be of the conventional type. However, where it is desirable to preserve the same type of construction throughout the filter it is possible to use evanescent sections at the end.

FIGS. 9(*a*), 9(*b*) and 9(*c*) show a six-section maximally flat direct-coupled filter with a 3 db. bandwidth of 37 mc./sec. The 3 db. limit frequencies are 3,983 mc./sec. and 4,020 mc./sec. the construction being in 2=2/3 in guide. There are evanescent mode coupling sections 1 between the resonant cavities 2. Each cavity 2 has a 2-BA coarse-tuning screw 3 and an 8-BA fine-tuning screw 4, and each evanescent section 1 has A 0-BA coupling adjustment screws 5. Inductive arises 6 at the end of the filter are conventional.

The physical dimensions of the filter are given below:

A—16.716 in., B—0.030 in., C—1.728 in., D—1.146 in., E—1.683 in., F—1.394 in., G—1.683 in., H—1.448 in., I—0.500 in., J—0.854 in.

The first step in designing the filter was to calculate the values of the prototype sections using Cohn's paper. For a maximally flat filter, using the same notation as Cohn, the values are:

$g_1=g_6=0.518$; $g_2=g_5=1.414$;

$g_3=g_4=1.93$

The susceptance connecting the i^{th} and $i+1^{th}$ cavities are given by:

$$B_{i,\,i+1}=\frac{\left(1-\frac{L^2}{g_i g_{i+1}}\right)\sqrt{g_i g_{i+1}}}{L} \qquad (12)$$

where

$$L=\pi\left[\frac{\lambda_{g1}-\lambda_{g2}}{\lambda_{g1}+\lambda_{g2}}\right] \qquad (13)$$

The guide wavelength terms λ_{g1} and λ_{g2} are the values occurring at the 3 db. band limits.

Taking $\lambda_{g1}=11.207$ cm. and $\lambda_{g2}=10.981$ cm. $L=0.0321$.

Substituting for L and g in (12) the values of $B_{i,\,i+1}$ are:

$$B_{12}=26.6;\ b_{23}=51.6;\ B_{34}=60.2.$$

The value of the outer susceptances B_{01} and B_{60} are obtained by substituting $g_o=L$ in (12) which gives

$$B_{01}=3.77=B_{60}.$$

The insertion loss of B_{12}, B_{13} and B_{34} are from (9):

$L_{12}=22.49$ db$=L_{56}$

$L_{23}=28.23$ db$=L_{45}$

$L_{34}=29.57$ db.

The dimensions of the evanescent guide section may now be calculated. Choosing a symmetrical junction with guide widths of 1 in. (evanescent section) and 2 in. (propagating section) respectively, the value $XZ_o/2a\lambda_g$ may be obtained from FIG. 6. For $a'/a=0.5$;$\lambda_g=11.1$ cm. $X(Z_o$ normalized) is given by $X=0.31$. From (11) the electrical "half-length" of the cavity is:

$$\theta/2=72.8°$$

or a total length, for identical junctions at each end of the cavity, of 145.6°. This gives a value of

$$l_c=(145.6\times 11.1/360)=4.49 \text{ cm. (or } 1.77 \text{ in.)}$$

Reducing this value by 5 percent to allow for tuning screws we have:

$$l'_c=1.682 \text{ in.}$$

The details of the inductive irises at the end of the filter remain to be settled. These are conventional symmetrical irises the dimensions of which may be obtained from any standard text.

The performance of the filter is shown in FIG. 11.

The field distribution in evanescent modes conveying full power through the section is of interest. FIG. 12 shows the measured field distribution in a typical three-section filter using evanescent couplings throughout. The filter was constructed with a slotted line section and measurements made with a probe (with very small insertion) in the usual way. When measuring the field in the various sections it was necessary to close up the slot in order to prevent radiation from the asymmetrical end sections. This was achieved by fitting a number of tongued sections (similar to the probe section) into the slot. These were linked to each other and attached to the probe so that they travelled along the guide with it.

FIG. 14 gives the calculated field along a transmission line of imaginary characteristic impedance which is terminated for full energy transfer (FIG. 13). The network is symmetrical and it is, therefore, sufficient to bisect the network and calculate E_1/E as a function of l (equation (8)). The values of γ and x_1 (x_1 is the equivalent reactance presented by the cavity at the line terminals) are chosen for a typical example. It will be seen that a standing wave exists on the line reaching zero in the dissipationless case at the center and comparatively small values if the loaded Q is high. The general character of results is the same at that shown in the experimental results of 12.

It is of interest that the magnitude of the electric field behaves in much the same way as if an iris of very high susceptance were located at the center of the intercavity coupling section. The end sections behave differently owing to their asymmetrical nature. The semiresonant rise in voltage predicted by eq. (8) can be clearly seen in FIG. 12.

The essential difference between this type of band-pass filter and conventional types lies in two features: the physical length of the coupling sections and the nature of the field existing within the sections. The insertion loss of an evanescent section depends both on its width and length and, therefore, the length may vary over a wide range of values. This means that the filter will be somewhat longer than the conventional direct-coupled type although, in general, it need not be longer than the quarter-wave coupled filter. This is a disadvantage where the shortest possible filter is desired. However, where a modest increase in length can be tolerated the filter has a number of advantages.

a. The freedom of choice in the length of the coupling sections means that the overall length of a given type of filter can be maintained irrespective of its center frequency. This follows, because of the longer wavelengths the greater cavity length is obtained by reducing the length (and if necessary, the width) of the evanescent section. Thus, only one fixed body size, with appropriate drillings, needs to be retained for a complete waveguide band. This simplified storage problems in production. The inserts, which form the evanescent sections, and are considerably more robust than conventional irises, are then the only variables.

b. The mechanical tolerances for large values of B, by whatever method that are obtained, are very severe. For example, if the coupling sections in the center of the filter of FIG. 9 are replaced by symmetrical inductive irises of the appropriate value, the tolerance on the width dimension (i.e., the gap between the irises) is only one-third of that which may be permitted with the corresponding evanescent guide dimension. Other types of obstacles, such as holes and multipost arrangements require even stricter tolerances. Therefore, filters in which the bandwidth is significantly narrower than about 1 percent (the above example) will require considerably less severe tolerances if evanescent coupling sections are employed.

c. The style of construction considered so far can be produced by inserting milled blocks in waveguide in order to

produce the evanescent sections. Another method is to spray metal on a suitable mandrel. The filter can be produced in two halves by milling from solid material. This method of construction can be expected to yield very high precision and has the advantage of eliminating soldered joints.

d. Variable bandwidth filters can be produced by using variable-cutoff evanescent coupling sections such as that shown in FIG. **10**, which incorporates a movable block **7** locked in any desired position by a screw **8** and additional to fixed blocks **9** and **10**.

e. The most important feature of this filter is its potential for eliminating the parasitic passbands that are characteristic of all band-pass filters at microwave frequencies. Normally, suppression of these passbands requires an additional filter which then affects the midband insertion loss and reflection. However, with evanescent coupling sections this disadvantage is avoidable. One or more of the coupling sections can be so designed that the frequency it is desired to suppress propagates in the coupling section instead of being an evanescent wave, like the desired passband. It is then possible to insert parasitic suppressors in the section concerned which will have a very large effect on a progressive wave, but none on an evanescent wave.

Conventional resonant-cavity waveguide band-pass filters suffer from unwanted transmission bands above the desired passband. These passbands occur for two reasons. Firstly, in direct-coupled resonant-cavity filters resonance occurs at frequencies for which the electrical length of the cavity is $n.\lambda g/2$ (where n is an integer and λg is guide wavelength). In addition to these multiple resonances, which are approximately in harmonic relationship, further resonances can be expected as a result of higher order modes propagating. In general, individual cavities tend to resonate at slightly different frequencies and interaction between these resonances occurs. As a result, multiple narrow passbands occur at unpredictable frequencies and, therefore, there is no way of knowing whether a filter will provide satisfactory harmonic rejection, for example, before it is built. If, in addition, a specific filter design is to be tuned over a wide range of frequencies the danger of harmonics penetrating the filter becomes great. In these circumstances, where suppression of high-power harmonics is essential, it is necessary to cascade a second filter giving broad band suppression. This may take the form of a corrugated (or waffle-iron) low-pass filter or a leaky wall filter. The former is based on the classical lumped circuit analogue and is a reflection filter; the second permits the harmonics to leak through the wall of the filter to auxiliary guides where they are dissipated. The difficulty with both of these filters is that they affect the passband performance adversely. Both filters add unavoidable dissipation and reflection that cannot be matched out (because the phase of the reflection varies rapidly with frequency).

The behavior of filters using evanescent mode couplings is completely different from conventional iris-type filters at frequencies will above type filters at frequencies well above the main transmission band. Above a critical frequency the coupling section propagates and the resonant cavities formed by the evanescent coupling sections virtually disappear. Instead, the filter behaves, practically, as a straight through device of nearly zero attenuation and therefore, only has minor internal reflections. In this form it is worse than the conventional filter, but the elimination of large internal reflections in the filter is a prerequisite to providing predictable suppression of unwanted frequencies. The second necessary condition is that the coupling waveguide should behave differently for the passband and unwanted frequencies, respectively. In this way devices which affect the unwanted frequency will have little or no effect on the desired frequency. The behavior of evanescent couplings, as described above, suits them to this application.

In general, the suppression of unwanted frequencies presents itself in two possible forms. One occurs when the filter functions in a closed circuit e.g., it provides filtering in the output of an oscillator. The unwanted frequencies, virtually entirely harmonics, are known. If the device is a high-power oscillator a very large amount of suppression may be necessary, but it will be required at specific frequencies i.e., narrow-band rejection circuits of very high rejection will be required. The second example occurs when the filter is located in an open-circuit system i.e., a system connected, for example, to a wide-band aerial. Signals over a wide band of frequencies will be likely to be received and, thus, wide-band suppression is necessary. In general, the degree of suppression required will be required to be less than the first example.

FIGS. **15**(*a*) and **15**(*b*) shows an evanescent coupling section **11** between resonant cavities **12** incorporating a narrow-band harmonic rejection filter. At the desired transmission band the coupling section is designed to be evanescent but the dimensions are chosen so that propagation occurs at the frequency to be suppressed. A three-section rejection filter is shown, the stubs **13** containing semiconductor tuning pistons **14** being conveniently connected to the guide by the broad-band resonant slots **15**. Approximately 35 db. rejection per section can be obtained in a typical example over a relatively narrow band. Thus, a three-section filter built in the coupling sections as shown gives rejection, at a given frequency, in excess of 100 db. If it is desirable to absorb the undesired harmonic, rather than reflect it back to the source, the filter can be converted into a constant resistance filter (British Pat. No. 1,018,923 G. F. Craven 7) and the energy absorbed into the matched load **16**, FIGS. **16**(*a*) and **16**(*b*) in which like references have been used as for FIG. **15**. The rejection of one section is largely lost but the advantage of being able to absorb the energy is often significant. The performance of a filter using one of the sections of FIG. **16** is shown in FIG. **17**. The midband insertion loss of this section (0.4 db.) compared with a section in which no rejection sections were included was identical within the limits of measurement. The reason for the absence of additional loss is that the guide section incorporating the load is beyond cutoff at the main transmission frequency and can easily be made sufficiently long to prevent measurable loss. In this respect, this form of the rejection filter has something in common with leaky wall filters which dissipate the energy by coupling to a large number of guides (which are beyond cutoff at the main transmission frequency). However, this filter avoids the main disadvantage of the leaky wall filter: the main transmission frequency propagates in the dominant mode and therefore, harmonics propagate in several modes. Multiple coupling to all possible modes is then necessary and this produces measurable insertion loss at the main frequency. In many applications the present filter is likely to be superior to the leaky wall filter, because the guide size of the coupling section may be chosen so that the harmonic to be suppressed propagates in the dominant mode only (the main frequency propagating in an evanescent mode). Thus, the present filter will be much simpler, have less insertion loss and combine the functions of band-pass filter and harmonics suppression filter.

Since the filter of FIG. **15** is essentially a reflection filter, only one suppression filter of this type is permissable per complete band-pass filter. This is because, if additional evanescent coupling sections are converted into suppression filters, mutual cancellation can be expected with further parasitic passbands resulting. Thus, the amount of suppression that can be obtained is limited to the number of suppression sections that can be contained in one coupling section.

However, this limitation does not apply if the suppression filters are dissipative. Ordinarily, dissipation in the main coupling arm is not acceptable, but if the suppression filter is designed as a constant resistance type a considerable resistive attenuation is obtainable at the parasitic frequency. At the desired passband frequency the coupling guide is evanescent and therefore no dissipation can occur.

Another form of narrow-band suppression filter is shown in FIG. **18**. Two resonant cavities **17** are coupled by an evanescent coupling section **18**. Resonant coupling slots **19** couple the evanescent section **18** to a length of circular

waveguide 20 terminated in a matched load 21 to absorb the energy. This type of suppression filter based on resonant slot hybrid junctions can give isolation of 20–40 db. over a band which is broader than the filter of FIG. 16. The attenuation characteristic of the filter of FIG. 18 is shown in FIG. 19.

A further form of absorption narrow band filter is shown in FIG. 20. Here, two resonant cavities 22 are coupled by an evanescent coupling section 23 in which is a cavity 24 resonant at the propagated parasitic frequency to transfer the power at the parasitic frequency via a coupling hole 25 into a second cavity 26 where it is absorbed by a matched load 27. A series rejection filter 28 provides isolation, in the form of a short circuit across the guide, from the remainder of the circuits in the complete filter. At the passband frequency neither of the cavities 24, 26 can resonate because the guide is in the evanescent condition.

The attenuation characteristic and equivalent circuit of the filter of FIG. 20 are shown in FIGS. 21 and 22 respectively.

A point to be noted in a filter incorporating this type of parasitic suppressor is that, if power at the spurious response is to be absorbed and not merely reflected, the filter is no longer a symmetrical device. Thus input and output ports have to be specified.

A further type of narrow-band rejection filter of the series resonance type is shown in FIGS. 23(*a*) and 23(*b*). Resonant cavities 28 are coupled by an evanescent coupling action 29 which contains a central short-circuiting screw 30 across the guide. Since this is a reflection filter, there can be only one per complete filter for the reasons already given.

The attenuation characteristic of the filter of FIG. 23 is shown in FIG. 24.

A further form of narrow-band absorption filter is shown in FIG. 25 which exploits the difference existing between the propagated mode of the parasitic frequency and the evanescent mode of the passband frequency. Two resonant cavities 31 are coupled by an evanescent coupling section 32 which contains a cavity 33 resonant at the parasitic frequency. There is a cavity tuning screw 34 and an iron dust slug 35. A series rejection filter 36 provides isolation, in the form a short circuit across the guide, from the remainder of the circuits in the filter.

As shown in FIG. 26, the amplitude of the electric field in the evanescent mode is at a minimum at the center of the coupling section. However, the amplitude of the electric field in the propagating mode (the harmonic) is at a maximum in the resonant cavity 33 built in the coupling section. The relative amplitudes are in the approximate ratios of Q_1, Q_2 and Q_1 and Q_2 are the loaded Q's of the main filter cavities 31. Thus an absorption screw (the slug 35) in the cavity 33 absorbs the harmonic energy without having any significant effect on the main transmission frequency.

Wide-band rejection filters are based on lumped circuit low-pass networks and take the form of corrugated waveguide filters, FIGS. 27(*a*) and 27(*b*) which show two resonant cavities 37 coupled by an evanescent coupling section 38 incorporating a corrugated low-pass filter 39. A later and more complex version is the waffle-iron filter. Such networks are characterized by rejection over a very wide band FIG. 29 and are employed in the following way. The low-frequency cutoff of the guide containing the suppression section is set at a much higher frequency than the other coupling sections in the filter. If this LF cutoff is suitably chosen this section will propagate freely in the band in which the other resonant cavities are providing high rejection, FIG. 30. No serious harm will result from this in a multisection filter (more than 3 cavities) for only the rejection of one cavity will be lost. However, at higher frequencies where the other coupling sections are propagating freely the suppression section (if its HF cutoff is correctly chosen) will be functioning as a low-pass filter and providing considerable rejection. In this way the usual spurious passbands can be suppressed over a wide band. A filter has been constructed confirming the effectiveness of this type of suppression section. As a general principle it is preferable to use only one suppression section per filter because of the danger of two sections interacting and causing a new parasitic passband. However, where more than one suppression section is necessary in order to provide the desired rejection, two suppression sections can be included as shown in FIG. 31(*a*) and 31(*b*) in which reflective interaction between the two suppression sections 38(*a*) and 38(*b*) is prevented by a matched load 40 terminating a stub 41 between the suppression sections absorbing the reflected energy from either filter.

It is to be understood that the foregoing description of specific examples of this invention is made by way of example only and is not to be considered as a limitation on its scope.

I claim:

1. A waveguide band-pass filter comprising:
at least one waveguide section which is evanescent for the passband frequency of the filter, said waveguide section being dimensioned to propagate parasitic passband frequencies;
at least a first and second resonator cavity filter section coupled together through said one evanescent waveguide section;
a narrow-band absorption filter coupled to said evanescent waveguide section for suppressing parasitic frequencies, said absorption filter is of the constant resistance type and includes:
at least two stubs;
means including at least two broadband resonant slots for coupling said at least two stubs to said evanescent waveguide section;
means coupled to one of said stubs for absorbing reflected energy from other of said stubs coupled to said evanescent waveguide section;
at least one tunable screw coupled to the other of said at least two stubs.

2. A waveguide band-pass filter comprising:
at least one waveguide section which is evanescent for the passband frequency of the filter, said waveguide section being dimensioned to propagate parasitic passband frequencies;
at least a first and second resonator cavity filter section coupled together through said one evanescent waveguide section;
a narrow-band absorption filter coupled to said evanescent waveguide section for suppressing parasitic frequencies, said absorption filter includes:
a third cavity resonant at the parasitic frequency;
a fourth cavity;
means including a coupling hole in each of said third and fourth cavities to transfer energy at the parasitic frequency from said third cavity to said fourth cavity;
a matched load coupled to said fourth cavity for absorbing said transfer energy; and
a series rejection filter section coupling said third and fourth cavity to said evanescent waveguide section.

3. A waveguide band-pass filter comprising:
at least one waveguide section which is evanescent for the passband frequency of the filter, said waveguide section being dimensioned to propagate parasitic passband frequencies;
at least a first and second resonator cavity filter section coupled together through said one evanescent waveguide section;
a narrow-band absorption filter coupled to said evanescent waveguide section for suppressing parasitic frequencies, said absorption filter includes:
a third cavity resonant at the parasitic frequency;
an obstruction mounted within said third cavity for absorbing the
parasitic frequency energy without having any significant effect on the passband energy; and
a series rejection filter coupling said parasitic frequency cavity to said evanescent waveguide section.

4. A filter according to claim 3 wherein said obstruction includes an absorption screw.