Design Considerations for High-Power Microwave Filters*

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Summary—The need for high-power filters is reviewed briefly, and various design approaches are discussed. The major portion of the paper treats the power-handling capacity of multiple-resonator filters using inductive windows or posts as coupling elements. A formula is derived that gives the relative power capacity of a waveguide filter of this type in terms of the bandwidth and cavity dimensions, and the element values of the low-pass prototype filter. By means of this formula it is shown quantitatively how high-power ratings may be achieved through the use of enlarged cavities. Methods for eliminating spurious filter responses and of reducing the reflected energy are discussed.

I. Introduction

In the microwave range, equipments often interfere with each other even when operating at different frequencies. This interchannel interference can arise from the generation and radiation of large amounts of RF energy outside of the operating bandwidth of a transmitter, or it can be the result of inadequate RF selectivity in a receiver. For example, magnetrons are known to produce strong harmonics, and are also likely to emit spurious "moding" energy at frequencies that are not harmonically related to the fundamental frequency. Some of these unwanted components may be at a level only 20 db below the fundamental, and hence are capable of causing serious interference to receivers in other systems. Fortunately, interchannel interference can be eliminated through the use of RF filters. High-power filters may be used between a generator and an antenna to prevent the radiation of signals outside the operating bandwidth, and low-power filters may be used at the input of a receiver to prevent the entrance of strong signals of undesired frequency. A discussion of filters in the high-power class is given in this paper. Low-power filters represent less of a problem, and considerable design information may be found in the literature.

In the design of filters intended for high-power applications, one must prevent electric field strengths approaching the breakdown point. Sharp edges in high-field-strength regions must therefore be avoided, and resonant buildup of the field must be limited to a safe value. One approach to the high-power filter problem has been made by Vogelman,[1] who has proposed a varying-impedance waveguide structure in which the changes of height are tapered rather than abrupt. Vogelman has shown how to compute the pass-band and spurious responses of this filter and has verified experimentally its frequency response and high-power capability. A second approach has been used by Wheeler and Bachman[2] in which resonant irises and resonant posts are quarter-wave coupled in a waveguide to synthesize an m-derived band-pass filter. Breakdown is avoided by evacuating the filter, which greatly increases the breakdown field strength. A third approach proposed by Torgow[3] uses a circuit of hybrid junctions, loads, and waveguides of different cutoff frequencies to create pass bands and absorbing bands. Since resonant elements are avoided, the power-handling capacity can be very high. A fourth approach is to use the multiple-cavity type of band-pass filter, in which inductive windows or posts are used for coupling the waveguide resonators. The power-handling capacity of this configuration has been studied recently by the author of this paper, and it is shown here that this filter can have virtually as large a power rating as desired, up to the full rating of the terminating waveguide.

II. Power-Handling Formulas for Multiple-Resonator Filters

A multiple-resonator band-pass filter can be designed[4] to have the response function of a lumped-constant low-pass prototype filter, where, through an appropriate transformation, zero frequency for the latter corresponds to the center of the pass band for the former (see Fig. 1). Thus, with the proper choice of coupling-element values and cavity lengths, the waveguide filter and prototype filter of Fig. 2 are equivalent. In this figure, $g_1, g_2, \cdots, g_i, \cdots, g_n$ are element values of the prototype filter (in farads for i odd and henries for i even), b_T is the height of the terminating guides, $b_1, b_2, \cdots, b_i, \cdots, b_n$ are cavity heights, $m_1, m_2, \cdots, m_i, \cdots, m_n$ are integers equal to the cavity lengths in half guide wavelengths, E_T is the maximum electric field strength in the matched output waveguide, and $E_1, E_2, \cdots, E_i, \cdots, E_n$ are the maximum electric field strengths in the cavities. In terms of these quantities, the cavity field strengths at the center of the pass band are given

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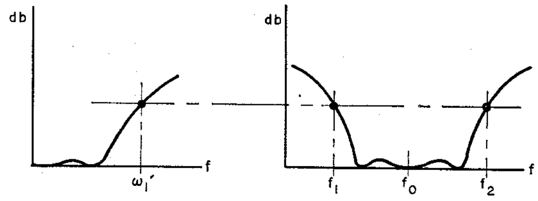
[1] J. H. Vogelman, "High-power microwave filters," 1958 IRE National Convention Record, pt. 1, pp. 84–90. Also, "High-power filters using higher-order-mode resonance," presented at PGMTT Natl. Symp., Stanford, Calif., May 7, 1958.

[2] H. A. Wheeler and H. L. Bachman, "Evacuated waveguide filter for suppressing spurious transmission from high-power S-band radar," this issue, p. 154.

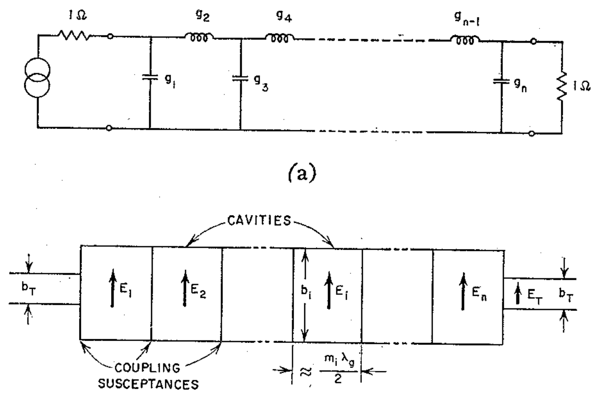
[3] E. N. Torgow, "Hybrid junction-cutoff waveguide filters," this issue, p. 163.

[4] S. B. Cohn, "Direct-coupled-resonator filters," Proc. IRE, vol. 45, pp. 187–196; February, 1957.



(a) (b)

Fig. 1—Correspondence between (a) low-pass prototype response and (b) equivalent band-pass response.



(b)

Fig. 2—Low-pass prototype filter and equivalent waveguide-cavity filter.

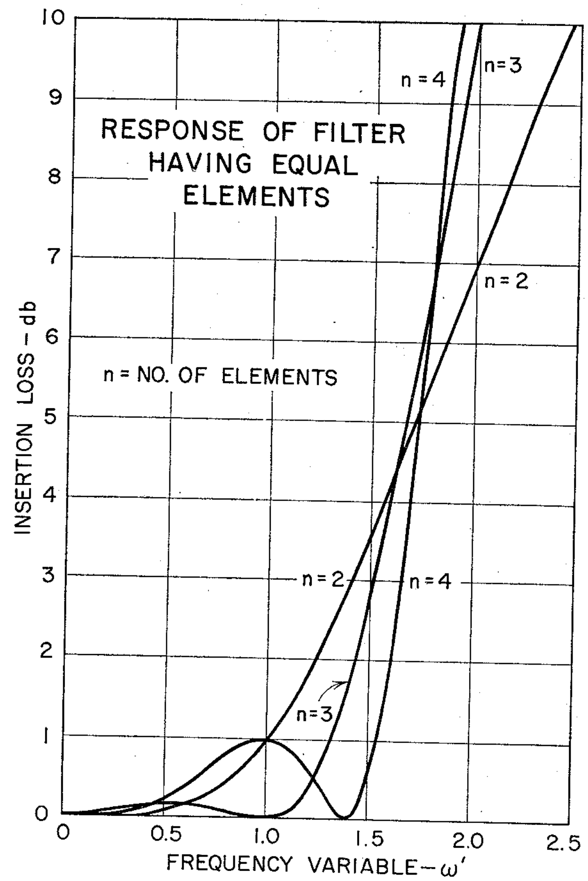


Fig. 3—Insertion-loss vs frequency response of multiple-cavity filter with $g_i=1$.

by the following formula, which is derived in Section V:[5]

$$E_i = E_T \frac{\lambda_0}{\lambda_{g0}} \sqrt{\frac{2g_i\omega_1' b_T}{m_i \pi w b_i}} \tag{1}$$

where $w=(f_2-f_1)/f_0$, and where $\omega_1'=2\pi f_1'$ for the prototype filter and f_1 and f_2 for the band-pass filter are corresponding points of equal insertion loss (see Fig. 1). Note that ω_1', f_1, and f_2 may be taken as desired either at the edges of the pass band or at any insertion-loss level in the stop band. In this formula it is assumed that the waveguide width and guide wavelength are the same throughout. [See (14) for generalization of (1).]

It is desirable in a high-power multiple-cavity filter to have equal electric field strengths in the cavities, since this condition maximizes the power-handling capacity for a given degree of selectivity. Examination of (1) shows that for cavities of equal size, the condition for equal electric field strength at the pass-band center is that $g_1, g_2, \cdots, g_n$ all be the same. The response function of the prototype low-pass filter for this special case is shown in Fig. 3 for various numbers of elements and with all values of g_i set equal to unity.[6] It is seen that only the central portion of the pass band of the equivalent band-pass filter can be used if the insertion loss (due to reflection) is to be low.

Fig. 4 shows how the electric field strengths vary with frequency for the case of $g_i=1$ in the prototype filter. These curves were computed from the prototype filter recognizing that the voltages across the capacitances and the currents through the inductances are proportional to the field strengths in the respective cavities. It is seen in Fig. 4 that the field strengths are equal only at $\omega'=0$, that is, at the center of the pass band of the equivalent band-pass filter. The power rating at any frequency with respect to the power capacity at band center is equal to the square of the reciprocal of the largest relative electric field strength at that frequency. Thus, for example, in the case of a four-cavity filter having $g_i=1$, the power capacity at $\omega'=0.4$ is 80 per cent of that at $\omega'=0$, and at $\omega'=1$ it is 50 per cent.

The power-handling capacity of the filter is defined best as a percentage of the power rating of the terminating waveguide. Thus, if the filter and terminating waveguide are under equal air pressure, the relative power capacity is as follows:

$$\text{relative power capacity} = \left(\frac{E_T}{E_i}\right)^2 \cdot 100 \text{ per cent}, \tag{2}$$

where E_i is the largest of the cavity field-strength values in the filter at the frequency in question. Hence,

[5] The portion of (1) under the square-root sign is actually dimensionless, since the units of $g_i\omega_1'$ are canceled by the prototype-filter load resistance that would appear in the expression if it had not been set equal to one ohm.

[6] G. L. Ragan, "Microwave Transmission Circuits," M.I.T. Rad. Lab. Ser., vol. 9, ch. 10, by A. W. Lawson and R. M. Fano, p. 681; McGraw-Hill Book Co., Inc., New York, N. Y.; 1948.

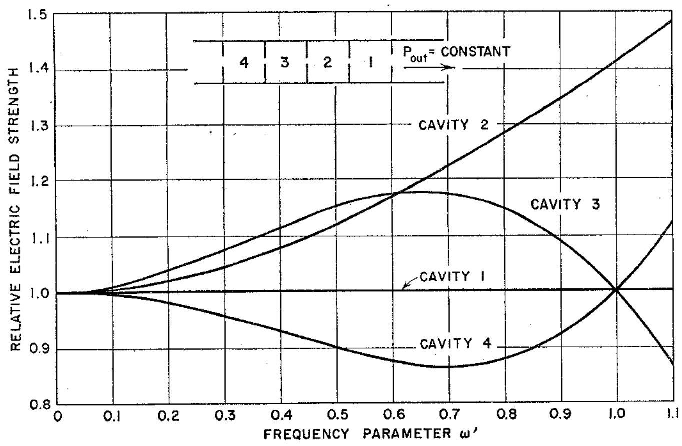


Fig. 4—Effect of frequency on electric field strengths in multiple-cavity filter with $g_i = 1$.

relative power capacity

$$= \frac{m_i \pi w b_i}{2 g_i \omega_1' b_T} \left(\frac{\lambda_{g0}}{\lambda_0} \right)^2 \cdot 100 \text{ per cent.} \quad (3)$$

The subscript i to be used in (3) is the number between 1 and n that minimizes this expression for a given n-resonator filter.

Examination of (1) shows that the cavity field strengths may be made equal even with unequal g_i values, if m_i and b_i are also made variable in the filter. Thus the superior maximally flat or equal-ripple response functions may be utilized without impairing the power rating of the filter, if the different cavities are properly proportioned. The larger cavities, of course, will have an increased number of undesired resonances, but the smaller cavities will tend to suppress these in the overall response.

III. Calculated Examples

A few examples will now be given of the relative power capacity of typical filters. As a first example, assuming $g_i = 1$, $\omega_1' = 1$, $w = 0.2$, $\lambda_0/\lambda_{g0} = 0.8$, $m = 2$, and $b_i = b_T$, (3) shows that at the center of the pass band the filter can be rated at 99 per cent of the power rating of the terminating waveguide. As a second example, assume that 40 db of insertion loss is required at the edges of a 20 per cent frequency band. For a four-cavity filter and $g_i = 1$, calculation shows that $\omega_1' = 3.89$ at the 40-db point of the prototype filter, and therefore $\omega_1' = 3.89$ and $w = 0.2$ are corresponding values. Now if we assume $\lambda_0/\lambda_{g0} = 0.8$, $m_i = 2$, and $b_i = 2b_T$, then the power rating at band center is 50.5 per cent of the power capacity of the terminating waveguide.

As a comparison to the last example, consider the power rating of a four-cavity filter whose g_i values are designed to yield a maximally flat response, with the 40-db points still at the edges of a 20 per cent frequency band. (The necessary formulas can be found in the literature.)[4] Calculation shows that $g_1 = g_4 = 0.765$, $g_2 = g_3 = 1.848$, and that the 40-db point occurs at $\omega_1' = 3.16$. In this case, the largest field strength occurs in the middle cavities, and, assuming the same values of λ_0/λ_{g0}, m_i, w, and b_i/b_T as in the last example, the filter rating at band center is equal to 33.5 per cent of the power rating of the terminating waveguide. This should be compared with 50.5 per cent in the preceding example.

Another interesting question is how the power rating of a multiple-resonator filter changes with the number of resonators when the stop-band bandwidth is held constant. The pass band bandwidth will increase with the number of resonators, approaching the stop-band bandwidth in the limit, and it is therefore clear that the power rating will also increase. A calculation has been carried out for the case of equal elements and the parameters of the second example with the following results:

Number of Resonators	ω_1' at 40-db Point	Relative Power Capacity (per cent)
1	200	1.0
2	14.16	13.9
3	5.86	33.5
4	3.89	50.5
5	3.10	63.3
6	2.70	72.7
100	2.00	98.1

The advantage of using a large number of resonators is clearly evident. However, considerations of size, adjustment difficulty, and dissipation loss will place a limit of, perhaps, six to ten resonators in a practical filter.

IV. System Considerations for High-Power Filters

It is evident from the preceding discussion that high-power filters are likely to have numerous undesired spurious responses. Fig. 5 shows how several filters in cascade may be used to eliminate spurious responses out to very high frequencies. The filter of smallest bandwidth, Filter 1, must have larger cavities than the others in order to carry the required power. As a result it will almost inevitably have a number of spurious responses in the range of interest. These spurious responses can be suppressed by one or more filters of greater bandwidth and more widely spaced spurious responses, as shown in the figure.

The high-power band-pass filter considered above provides stop-band insertion loss through reflection of the incident energy. In some instances, this reflection may have an adverse effect on the power source, and the energy should be dissipated rather than reflected. As shown by the examples in Fig. 6, this may be done by means of a broad-band ferrite isolator, or by a circuit of hybrid junctions and a pair of identical band-pass filters so arranged as to divert the energy reflected from the filters into an auxiliary resistive load. Equivalent performance can be achieved by directional filters, which have the combined properties of directional couplers and of filters.[7] Directional filters would be particularly useful

[7] S. B. Cohn and F. S. Coale, "Directional channel-separation filters," Proc. IRE, vol. 44, pp. 1018–1024; August, 1956.

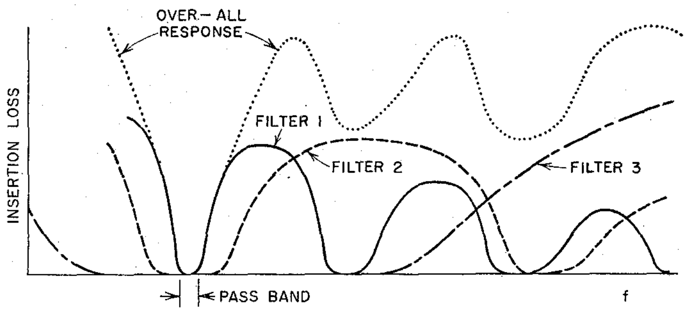
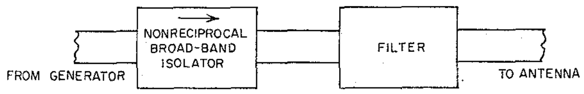
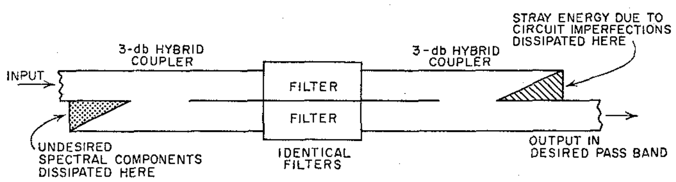


Fig. 5—Connection of filters in cascade to eliminate spurious responses.



(a)



(b)

Fig. 6—Suggested means for dissipating energy reflected from filter. Network in (a) utilizes a broad-band nonreciprocal isolator; network in (b) utilizes a pair of identical filters and hybrid junctions.

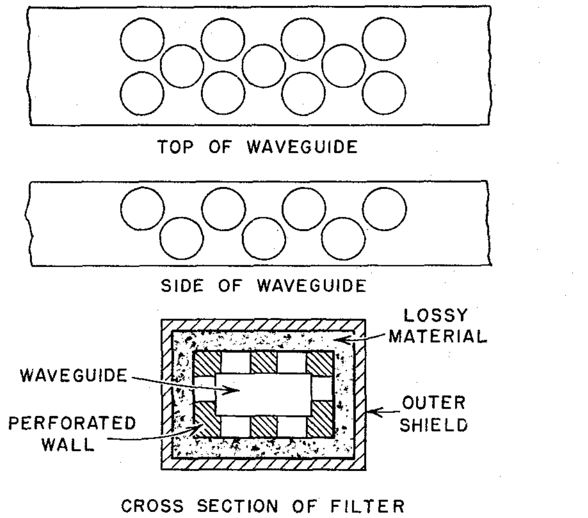


Fig. 7—Leaky-waveguide technique for dissipating unwanted energy above the desired signal frequency.

for removing discrete spurious frequencies and dissipating their energy into a load.

It is believed that wide-bandwidth high-power filters can also be achieved by leaky-wall structures like that of Fig. 7. Near the desired operating frequency the holes would be waveguides below cutoff, and very little energy would leak through them. Above the cutoff frequency of the holes, the energy would pass through the holes and be absorbed in the surrounding lossy material. By having a sufficient number of holes on all four walls of the waveguide, it should be possible to obtain a stop band free of spurious responses in any waveguide mode up to a very high frequency. Of course the edges of the holes in the broad walls should be rounded to maximize the power capacity.

V. Derivation of Field-Strength Formula for Multiple-Resonator Filter

The following derivation is based on a previous analysis of multiple-resonator filters.[4] In that analysis, the low-pass prototype filter of Fig. 2(a) was related to the waveguide filter of Fig. 2(b) by means of the intermediary transmission-line circuit of Fig. 8. The equivalence of Fig. 8 to the waveguide filter is a direct one, with the lines of characteristic impedance Z_0 and electrical length 180 degrees representing the cavity resonators at resonance, and the lines of characteristic impedance $K_{i,i+1}$ and electrical length 90 degrees representing the coupling susceptances at the proper reference planes. It has been shown[4] that the latter length is independent of frequency to the first order, and therefore the lines of characteristic impedance $K_{i,i+1}$ act as impedance inverting transformers over a wide band of frequencies. The equivalence of Fig. 8 to the low-pass prototype filter depends upon the change in frequency variable shown in Fig. 1.

For convenience in the analysis, the cavities and terminating lines are assumed to have equal characteristic impedance Z_0, and the cavity electrical lengths at resonance are assumed to be 180 degrees. At a later point in the analysis these quantities will be generalized. Also, for convenience the various elements are numbered from the load end of the filter.

Fig. 9 shows how the voltages and currents in the filter may be computed at the center frequency in terms of the voltage, V_T, across the load resistance. The phases of the various quantities are unimportant in this problem, and therefore only magnitudes of the voltages and currents are shown. The following simple relations between these magnitudes were used in the computation: for a 90-degree line of characteristic impedance Z_c,

$$V_{\text{in}} = Z_c I_{\text{out}} \quad \text{and} \quad I_{\text{in}} = \frac{V_{\text{out}}}{Z_c}; \tag{4}$$

and for a 180-degree line

$$V_{\text{in}} = V_{\text{out}} \quad \text{and} \quad I_{\text{in}} = I_{\text{out}}. \tag{5}$$

Starting at the load end and proceeding back through the filter, one can verify the expressions in Fig. 9, and show that the voltages at the centers of the resonators are of the following form:

$$V_1 = \frac{Z_0 V_T}{K_{01}} \qquad V_2 = \frac{K_{01} V_T}{K_{12}}$$

$$V_3 = \frac{Z_0 K_{12} V_T}{K_{01} K_{23}} \qquad V_4 = \frac{K_{01} K_{23} V_T}{K_{12} K_{34}} \quad \text{etc.} \tag{6}$$

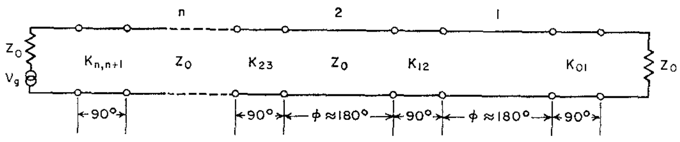


Fig. 8—Equivalent circuit of a multiple-resonator filter.

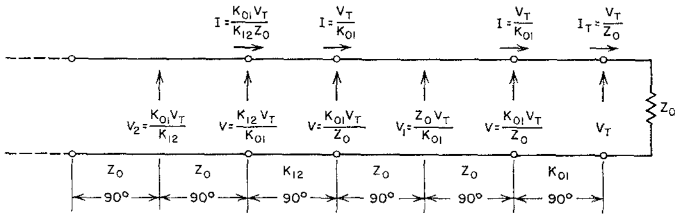


Fig. 9—Voltage and current magnitudes in the equivalent multiple-resonator filter.

In a waveguide filter the characteristic impedances $K_{i,i+1}$ are less than Z_0 and it is easily verified that the voltages V_i correspond to the maximum voltages in the various cavities.

The characteristic impedances $K_{i,i+1}$ have been shown[4] to be related to the element values g_i of the prototype filter by

$$\frac{K_{i,i+1}}{Z_0} = \frac{L}{\sqrt{g_i g_{i+1}}}, \qquad i = 1 \text{ to } n-1, \tag{7}$$

and

$$\frac{K_{01}}{Z_0} = \sqrt{\frac{L}{g_1}}\,. \tag{8}$$

The formula for $K_{n,n+1}$ is not as simple in general, but in the usual case of a symmetrical waveguide filter $K_{n,n+1} = K_{01}$. (If an unsymmetrical waveguide filter is of interest, Ref. 4 should be consulted for the correct expression for $K_{n,n+1}$.) Subject to the assumption of half-wavelength cavities of height and width equal to that of the terminating waveguide, the parameter L is given by

$$L = \frac{\pi w}{2\omega_i}\left(\frac{\lambda_{g0}}{\lambda_0}\right)^2 \tag{9}$$

where w is the relative bandwidth $(f_2 - f_1)/f_0$, and f_0, f_1, f_2, and ω_1' are defined in Fig. 1. λ_{g0} and λ_0 are the guide wavelength and free-space wavelength at the center frequency. (A more accurate form of (9) has been given,[4] but (9) is sufficiently accurate for power-capacity calculations.) At this point the limitation on cavity length and height may be removed. A study of the original derivation of (9) shows that

$$L_i = \frac{m_i \pi w b_i}{2\omega_1' b_T}\left(\frac{\lambda_{g0}}{\lambda_0}\right)^2 \tag{10}$$

where m_i is the length of the ith cavity in half guide wavelengths, b_i is the height of the cavity, and b_T is the height of the terminating waveguide. Combination of the above formulas results in

$$V_i = V_T\sqrt{\frac{g_i}{L}} = V_T\frac{\lambda_{g0}}{\lambda_0}\sqrt{\frac{2\omega_1' g_i b_i}{m_i \pi w b_T}}, \quad i = 1 \text{ to } n. \tag{11}$$

The voltages are related to the field strengths by

$$V_i = E_i b_i \quad\text{and}\quad V_T = E_T b_T, \tag{12}$$

and hence

$$E_i = E_T\frac{\lambda_0}{\lambda_{g0}}\sqrt{\frac{2g_i\omega_1' b_T}{m_i \pi w b_i}} \tag{13}$$

where E_T is the maximum electric field strength in the output waveguide, and E_i is the maximum electric field strength in the ith cavity.

Eq. (13) assumes that the cavity widths and terminating-waveguide widths are all equal. If this restriction is removed, one can show that (13) becomes

$$E_i = E_T\left[\frac{2g_i\omega_i' a_T b_T}{m_i \pi w a_i b_i}\right]^{1/2}\left[1-\left(\frac{\lambda_0}{2a_T}\right)^2\right]^{1/4} \cdot\left[1-\left(\frac{\lambda_0}{2a_i}\right)^2\right]^{1/4} \tag{14}$$

where a_i is the width of the ith cavity and a_T the width of the terminating waveguide.

VI. Conclusions

The coupled-resonator filters described in this paper are suited for use with high-power microwave tubes to suppress harmonics and spurious "moding" energy. The design techniques are believed to be sufficiently rigorous for practical structures, but are as yet unsupported by experimental verification. Further study of spurious pass bands and their elimination is desirable, and coupling configurations having optimum power capacity should be investigated. It is hoped that, despite their preliminary nature, the findings in this paper will prove of value to the engineer working in this relatively new field of high-power filtering.