Oct. 13, 1964 H. J. RIBLET 3,153,208

WAVEGUIDE FILTER HAVING NONIDENTICAL SECTIONS RESONANT AT SAME FUNDAMENTAL FREQUENCY AND DIFFERENT HARMONIC FREQUENCIES

Filed May 6, 1960 2 Sheets-Sheet 1

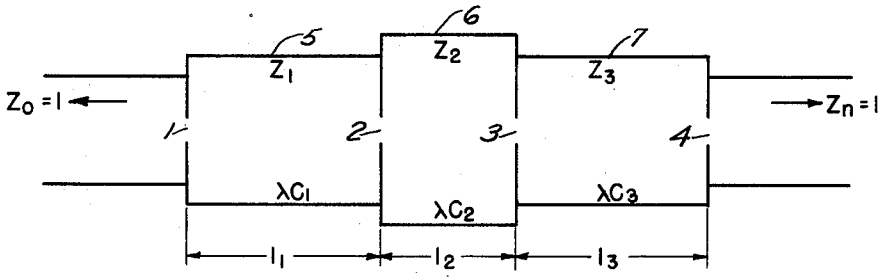


FIG. 1

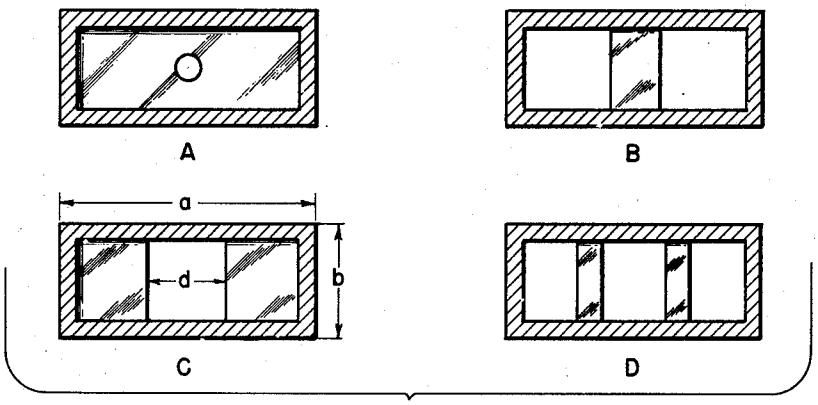


FIG. 2

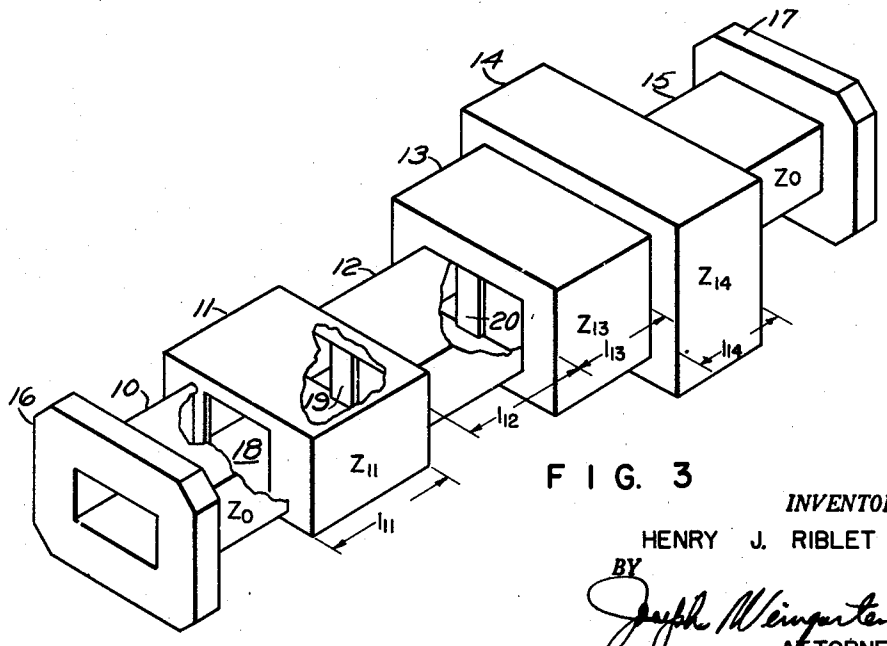


FIG. 3

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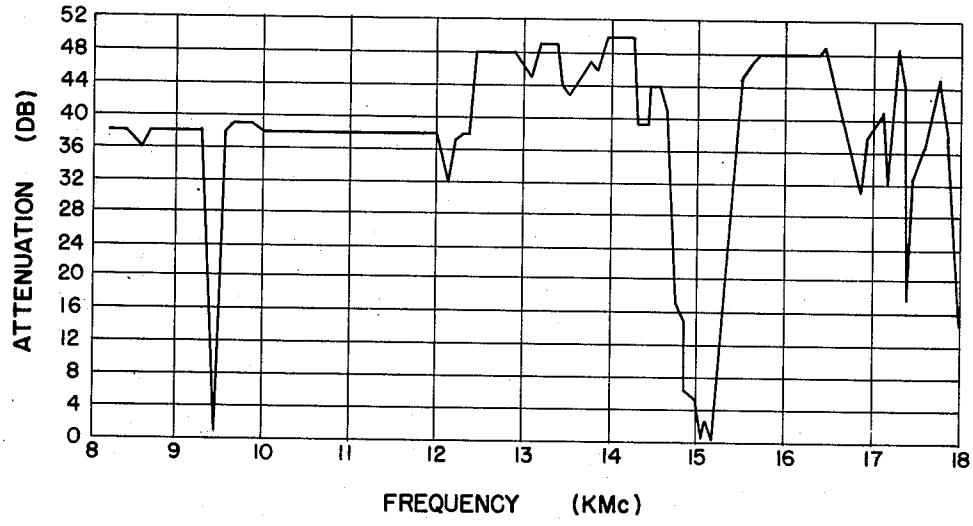


FIG. 4

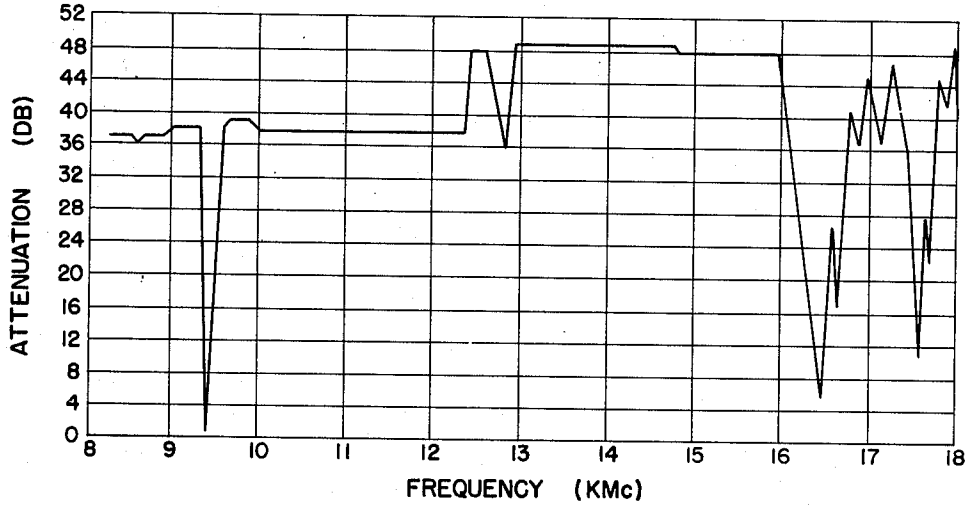


FIG. 5

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WAVEGUIDE FILTER HAVING NONIDENTICAL SECTIONS RESONANT AT SAME FUNDAMENTAL FREQUENCY AND DIFFERENT HARMONIC FREQUENCIES

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This invention relates in general to microwave filters employing sections of waveguide and more particularly pertains to a novel high-Q narrow band microwave filter in which all the waveguide sections are resonant at a single fundamental frequency but are not simultaneously resonant at any higher frequency.

It is well known in the microwave filter art that many of the techniques developed in connection with the synthesis of lumped parameter linear passive networks may be adapted for synthesizing microwave filters formed by a cascade of large, generally lossless, reflecting elements regularly spaced in a uniform waveguide. General synthesis procedures for "quarter wave coupled" and "direct coupled" filters are set forth in "Microwave Transmission Circuits," volume 9, M.I.T. Radiation Laboratory Series, pages 661–706. For both types of filters, synthesis procedures are described based on the use of a ladder network prototype having a prescribed insertion loss function. A closely related synthesis procedure for the design of direct coupled filters is also disclosed in a paper by S. B. Cohn entitled "Direct Coupled Resonator Filters" in the "Proceedings of the IRE," February 1957, volume 45, pages 187–196.

Generally, prior high-Q filters have consisted of a number of reflecting elements spaced within a waveguide of uniform cross section by one-half the wavelength in the guide of microwave energy at the center frequency of the filter. Because all the cavities formed in the uniform waveguide between reflecting elements are equal to or are multiples of $\lambda_g/2$ in length (λ_g, being the wavelength in the guide), all the cavities resonate at the same fundamental frequency and at all higher frequencies at which a single one of the cavities is resonant. In most uses of resonator filters, it is distinctly undesirable to have the filter pass all the higher frequencies at which the cavities are resonant. More commonly such filters are employed to pass only a continuous band of frequencies centered about the fundamental frequency. Accordingly, the present invention contemplates and has as a primary object the construction of high-Q waveguide filter in which the cavities of the filter are all resonant at a single fundamental frequency but are not simultaneously resonant at any higher frequency.

Another object of the invention is to provide a high-Q waveguide filter having a plurality of cavities, the design of the filter being such that individual cavities can be altered to optimise high power and insertion loss characteristics.

In accordance with the invention, a high-Q waveguide filter is constructed of direct coupled waveguide sections which in general have different cut-off frequencies and characteristic impedances (viz., a non-uniform waveguide), and reactive elements arranged in the sections in a manner such that each section is resonant at the same fundamental frequency. As the guide wavelengths are different in the different sections, the sections are not all simultaneously resonant at any higher frequency.

The construction of an embodiment of the invention and its manner of operation can be better understood by a perusal of the following exposition when considered in conjunction with the accompanying drawings in which:

FIG. 1 is a diagrammatic representation of a direct coupled filter employing reflective discontinuities;

FIG. 2 depicts various types of reactive impedance elements;

FIG. 3 is a perspective view of a filter constructed according to the invention;

FIG. 4 is a graph showing the frequency response characteristic of a conventional direct-coupled resonant cavity filter;

FIG. 5 is a graph showing the improved frequency response characteristic of a direct-coupled resonant cavity filter embodying the invention.

In a monograph entitled "A Unified Discussion of High Q Waveguide Filter Design Theory," by Henry J. Riblet, printed in the IRE Transactions of Microwave Theory and Techniques, vol. MTT–6, Number 4, October 1958, incorporated herein by reference, there is discussed the general design of filters which consist of a cascade of lossless reflecting elements spaced in a regular manner in a uniform waveguide. The following theoretical exposition, which is an extension of the theory developed in that monograph, explains the manner in which the design of a filter having non-uniform waveguide sections can be readily obtained once the design of the conventional uniform waveguide filter is available.

Consider the cascade of reflecting discontinuities separated by lengths of transmission line of different characteristic impedances and cut-off wavelengths shown in FIG. 1. The reflecting discontinuities are indicated as irises **1**, **2**, **3**, **4** and the transmission line as a waveguide partitioned by the irises into sections **5**, **6**, **7**. The waveguide sections each have a different characteristic impedance Z_1, Z_2, or Z_3 and a different cut-off wavelength λ_{c1}, λ_{c2}, or λ_{c3}. Further, it is assumed that the reflections from the irises are largely reactive. Now, if suitable resonant line lengths l_1, l_2, l_3, which in general are different, are associated with the reactive reflecting elements **1**, **2**, **3**, **4**, then the elements can be replaced, for a single frequency at least, by admittance inverters. The matrix representation of the network then is

$$\begin{bmatrix} 0 & j\sqrt{r_1} \\ \frac{j}{\sqrt{r_1}} & \end{bmatrix} \begin{bmatrix} \cos\theta_1 & j\sin\theta_1 Z_1 \\ \frac{j\sin\theta_2}{Z_1} & \cos\theta_1 \end{bmatrix} \begin{bmatrix} 0 & j\sqrt{r_2} \\ \frac{j}{\sqrt{r_2}} & 0 \end{bmatrix}$$

$$\begin{bmatrix} \cos\theta_2 & j\sin\theta_2 Z_2 \\ \frac{j\sin\theta_2}{Z_2} & \cos\theta_2 \end{bmatrix} \cdots \begin{bmatrix} 0 & j\sqrt{r_{n+1}} \\ \frac{j}{\sqrt{r_{n+1}}} & 0 \end{bmatrix}$$

where

$$\theta_i = \frac{2\pi l_i}{\lambda_{gi}}$$

in which

l_i is the section line length

λ_{gi} is the wavelength in the guide section for each frequency in the pass band under consideration

For narrow band widths this approaches

$$\begin{bmatrix} 0 & j\sqrt{r_1} \\ \frac{j}{\sqrt{r_1}} & 0 \end{bmatrix} \begin{bmatrix} 1 & 0 \\ \frac{j\sin\theta_1}{Z_1} & 1 \end{bmatrix} \begin{bmatrix} 0 & j\sqrt{r_2} \\ \frac{j}{\sqrt{r_2}} & 0 \end{bmatrix} \begin{bmatrix} 1 & 0 \\ \frac{j\sin\theta_2}{Z_2} & 1 \end{bmatrix}$$

$$\begin{bmatrix} 0 & j\sqrt{r_{n+1}} \\ \frac{j}{\sqrt{r_{n+1}}} & 0 \end{bmatrix}$$

Now for each $\sin\theta_1$, there is a t_i such that

$$\omega = t_1 \sin\theta_1$$

where ω is a normalized frequency variable which for all i's is given by the above equation. The expression for

t_1 is derived from defining Equation 24 given on page 364 of the monograph, i.e.,

$$t=\omega_0 \csc \frac{\pi(\lambda_{g_1}-\lambda_{g_2})}{\lambda_{g_1}+\lambda_{g_2}}$$

(where λ_{g_1} and λ_{g_2} are the guide wavelengths at which the skirt insertion loss is specified) by rewriting that equation as

$$t_1=\omega \csc \frac{\pi(\lambda_{g_{1i}}-\lambda_{g_{2i}})}{\lambda_{g_{1i}}+\lambda_{g_{2i}}}$$

where $\lambda_{g_{1i}}$ and $\lambda_{g_{2i}}$ are the guide wavelengths in each section at which the skirt insertion loss for the entire filter is specified, those guide wavelengths, in general, being different for each section because the sections are of different cross-sections.

The matrix product can then be rewritten as

$$\begin{bmatrix} 0 & j\sqrt{r_1} \\ \frac{j}{\sqrt{r_1}} & 0 \end{bmatrix} \begin{bmatrix} 1 & 0 \\ \frac{j\omega}{t_1 Z_1} & 1 \end{bmatrix} \begin{bmatrix} 0 & j\sqrt{r_2} \\ \frac{j}{\sqrt{r_2}} & 0 \end{bmatrix} \begin{bmatrix} 1 \\ \frac{j\omega}{t_2 Z_2} \end{bmatrix} \cdots \begin{bmatrix} 0 & j\sqrt{r_{n+1}} \\ \frac{j}{\sqrt{r_{n+1}}} & 0 \end{bmatrix}$$

and this can be written in ladder network form as

$$\begin{bmatrix} 0 & j\sqrt{\frac{r_1}{t_1 Z_1}} \\ j\sqrt{\frac{t_1 Z_1}{r_1}} & 0 \end{bmatrix} \begin{bmatrix} 1 & 0 \\ j\omega & 1 \end{bmatrix} \begin{bmatrix} 0 & j\sqrt{\frac{r_2}{t_1 t_2 Z_1 Z_2}} \\ j\sqrt{\frac{t_1 t_2 Z_1 Z_2}{r_2}} & 0 \end{bmatrix} \begin{bmatrix} 1 & 0 \\ j\omega & 1 \end{bmatrix} \cdots \begin{bmatrix} 0 & j\sqrt{\frac{r_{n+1}}{t_n Z_n}} \\ j\sqrt{\frac{t_n Z_n}{r_{n+1}}} & 0 \end{bmatrix}$$

Now

$$\frac{r_1}{t_1 Z_1} \frac{r_2}{t_1 t_2 Z_1 Z_2} \frac{r_3}{t_2 t_3 Z_2 Z_3} \frac{r_{n+1}}{t_n Z_n}$$

are known from classical synthesis procedures and are given by $\rho_1, \rho_2, \rho_3, \ldots \rho_n$. Therefore,

$$\frac{r_1}{Z_1}=t_1\rho_1 \quad \frac{r_2}{Z_1 Z_2}=t_1 t_2 \rho_2 \quad \frac{r_3}{Z_2 Z_3}=t_2 t_3 \rho_3 \cdots \frac{r_{n+1}}{Z_n}=t_n \rho_n$$

Moreover

$$\frac{r_1}{Z_1} \frac{r_2}{Z_1 Z_2} \frac{r_3}{Z_2 Z_3} \cdots \frac{r_{n+1}}{Z_n}$$

are precisely the voltage standing wave ratios introduced by the reflecting elements in the matched lines of characteristic impedance between which they are placed, assuming all impedances are normalized to the input and output transmission lines. The same design procedure used for the uniform line in the monograph can be used for lines of differing characteristic impedance if ρ is defined not as the inversion factor, but as the voltage standing wave ratio of the isolated reflecting elements and the t's in the general formulas for direct-coupled filters given on page 363 of the monograph are replaced by the t_i's associated with the neighboring waveguide section.

While capacitive reflective elements can be used, it is preferred that inductive irises be employed as coupling elements between the waveguide sections due to their better power handling capabilities, relative ease of fabrication, and lesser mechanical criticality. Some practical forms which the inductance may take are depicted in FIG. 2. The round hole inductive iris of FIG. 2A affords the highest unloaded Q for a resonant cavity, the center post inductive iris shown in FIG. 2B offers the lowest unloaded Q, while the symmetrical vane iris (FIG. 2C) and the symmetrical post inductance afford intermediate values of unloaded Q. The selection of one of those types in preference to the others for a given application is generally based upon a compromise between ease of fabrication and acceptable performance of the filter.

Referring now to FIG. 3, there is depicted a perspective view of a filter embodying the invention in which portions of the waveguide have been broken away to show some of the reflective irises. The filter comprises rectangular waveguide sections **10, 11, 12, 13, 14, 15**, adjacent waveguide sections being directly coupled. Input waveguide section **10** and output section **15** are identical in cross-section and each of those sections is provided with a flange **16, 17** for coupling to the line in which the filter is to be inserted. The waveguide sections **11, 12, 13, 14** are of varied cross-sectional dimensions and of different lengths. The sections **10** and **15** have the same characteristic impedance, designated Z_0, and the characteristic impedance of the sections **11, 12, 13, 14** are designated by the symbols Z_{11}, Z_{12}, Z_{13}, Z_{14}, respectively. The sections **11, 12, 13, 14** are of lengths l_{11}, l_{12}, l_{13}, l_{14} such that those sections are resonant at the center frequency of the band intended to be passed by the filter. The terms "center frequency" and "fundamental frequency" as used herein being synonymous. The input section **10** is coupled to the section **11** by an inductive iris **18** of the symmetrical vane type illustrated in FIG. 2A. The iris size is empirically chosen to produce a voltage standing wave ratio in one of the sections **11** in accordance with the calculated value for ρ_1 when the other section is terminated in its characteristic impedance. For example assuming the input signal to be directed into the flanged end of the section **10** and the section **11** to be terminated in its characteristic impedance, then the iris **18** must be of such dimensions as to cause the voltage standing wave ratio measured in the section **10** to yield the value determined for ρ_1. As the value of ρ_1 establishes the amount of energy coupled through the iris **18** from one section to the other and since the coupling action of the iris is reciprocal, the section **10** may be terminated in its characteristic impedance, the signal may be fed in through section **11**, and the voltage standing wave ratio may be measured in section **11**. After the dimensions of iris **18** have been ascertained, the same procedure is used to ascertain the size of the next iris, viz., the iris **19** coupling the section **11** to section **12**. Assuming the iris **19** to be of the center post type shown in FIG. 2B, the dimensions of that iris are chosen so as to produce a voltage standing wave ratio in one of the sections **11** or **12** in accordance with the calculated value for ρ_2 when the other of those two sections is terminated in its characteristic impedance. The size of the iris **20** at the next junction and of the other irises in the filter are determined in the same manner using the calculated values of ρ_3, ρ_4, etc.

It is evident that a procedure for designing a direct-coupled waveguide filter is here described in which each junction between two sections of waveguides of different cross-sections is independently examined in determining the dimensions of the iris coupling the two sections. The procedure, in essence, is not much more complicated than the procedure for designing a direct-coupled cavity filter of the uniform waveguide type. The improvement in performance which can be obtained with a filter of the type shown in FIG. 3 and the conventional uniform waveguide type filter is shown by a comparison of the graphs of FIGS. 4 and 5. Referring first to the graph of FIG. 4 depicting the performance of a conventional direct-coupled resonant cavity filter designed to pass a narrow band of frequencies in the vicinity of 9.5 kilomegacycles (kmc.), it can be seen that the filter also passes a relative wide band of higher frequencies lying within the range of 15.05 to 15.20, considering any frequency attenuated less than 3 db to be passed by the filter. In contrast, the FIG. 5 graph, depicted the performance of a direct-coupled filter constructed in accordance with the invention and designed to pass the same frequencies as the conventional filter, shows all higher frequencies are attenuated at least 6 db and that the range of higher frequencies at-

tenuated less than 20 db is substantially smaller than the comparable range for the conventional filter.

While the embodiment of the invention illustrated in FIG. 3 is fabricated from sections of rectangular waveguide, some reduction in insertion loss can be realized where circular waveguide sections are employed in the filter. In general, the insertion loss at resonance of a cavity filter decreases with the use of higher modes. It is known that even the lowest order circular waveguide mode (TE_{111}) gives insertion losses which are lower than can be obtained with the best rectangular waveguide cavity. This results from the extremely high unloaded Q's obtained in right circular cylindrical cavities. Where a cylindrical section is employed in the filter as a resonant cavity, it is treated in the same manner as a rectangular waveguide resonant cavity.

What is claimed is:

1. In a microwave filter of the type having a plurality of directly coupled waveguide sections and reflective reactive elements disposed in the waveguide causing each section to be resonant at the same fundamental frequency,

the improvement of an arrangement in which the sections are not all simultaneously resonant at frequencies higher than the fundamental, the improvement residing in employing directly coupled waveguide sections that have different cut-off frequencies and differerent characteristic impedances, and the length of each section being an integral half wavelength of wave energy in that section vibrating at the fundamental frequency whereby different sections are of different physical lengths.

2. In a microwave filter of the type having a plurality of directly coupled waveguide sections and reflective reactive elements disposed in the waveguide causing each section to be resonant at the same fundamental frequency,

the improvement of an arrangement in which the sections are not all simultaneously resonant at frequencies above the fundamental, the improvement residing in employing as the directly coupled waveguide sections contiguous sections that have different characteristic impedances and different cut-off frequencies, and the length of each section being an integral half wavelength of wave energy in that section vibrating at the fundamental frequency whereby differing sections are of different physical lengths.

3. In a microwave filter of the type having a plurality of directly coupled waveguide sections and elements disposed in the waveguide forming reflecting reactive irises which cause each section to be resonant at the same fundamental frequency,

the improvement of an arrangement in which the sections are not all simultaneously resonant at frequencies higher than the fundamental, the improvement residing in employing directly coupled waveguide sections that have different cut-off frequencies and different characteristic impedances, the iris forming elements being at the junctions of the sections, and the length of each section being an integral half wavelength of wave energy in that section vibrating at the fundamental frequency, whereby differing sections are of different physical lengths.

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