she is presently an Associate Professor. Her research interests are in the areas of microwave circuits, optimization, and computer-aided design.

Dr. Zaki is a member of Tau Beta Pi.

**Ali E. Atia** (M'69–SM'78) received the B.Sc. degree from Ain Shams University, Cairo, Egypt in 1962, the M.S. and Ph.D. degrees from the University of California at Berkeley in 1966 and 1969, respectively, all in electrical engineering. From 1962 to 1964, he was a Lecturer in the Department of Electrical Engineering, Ain Shams University. From 1965 to 1969, he held the positions of Research Assistant and Assistant Professor in the Department of Electrical Engineering and Computer Science, University of California, Berkeley. He joined COMSAT Laboratories in 1969 where he was engaged in research and development activities on antennas and various microwave components and subsystems for satellite transponders. Currently, he is Senior Director for Satellite System Design for INTELSAT Technical Services, COMSAT's World System Division, where he directs system design and evaluation activities of current and future satellite systems for use by INTELSAT. Dr.

Dr. Atia is a member of Sigma Xi and of AIAA.

# New Narrow-Band Dual-Mode Bandstop Waveguide Filters

JING-REN QIAN AND WEI-CHEN ZHUANG

*Abstract* —**A complementary relation between a dual-mode bandpass and bandstop waveguide filter is found. Then a new idea for constructing a bandstop filter is developed. Two trial samples of bandstop filters are constructed to demonstrate the principle.**

## NOMENCLATURE

| | |
|---|---|
| $M_{ij}$ | Normalized coupling coefficient between the $i$th and the $j$th loop. |
| $e_1$ | Equivalent source in the first loop. |
| $Z$ | Impedance of each loop. |
| $R, R_n$ | Equivalent loads in the first and last loop, respectively. |
| $n$ | Number of the loops in Fig. 1. |
| $i_k$ | Loop current in the $k$th loop. |
| $\boldsymbol{E}$ | Voltage matrix for a bandpass filter circuit. |
| $\boldsymbol{Z}$ | Impedance matrix for a bandpass filter circuit. |
| $\boldsymbol{I}$ | Current matrix for a bandpass filter circuit. |
| $R_0$ | Resistance looking into the source, characteristic impedance. |
| $\omega$ | Angular frequency. |
| $e_0$ | Source voltage. |
| $\boldsymbol{S}$ | A diagonal matrix. |
| $\boldsymbol{M}$ | Coupling matrix. |
| $M'_{01}, M'_{n\,n+1}$ | Normalized coupling coefficient between the main waveguide and the first loop and last loop, respectively. |
| $m$ | Turn ratio of the ideal transformer. |
| $i'_k$ | Loop current in the $k$th loop for a bandstop filter circuit. |
| $e'_1, e'_n$ | Equivalent source in the first and last loop for a bandstop filter circuit, respectively. |
| $R', R'_n$ | Equivalent load in the first and last loop for a bandstop filter circuit, respectively. |
| $\boldsymbol{E}'$ | Voltage matrix for a bandstop filter circuit. |
| $\boldsymbol{Z}'$ | Impedance matrix for a bandstop filter circuit. |
| $\boldsymbol{I}'$ | Current matrix for a bandstop filter circuit. |
| $\boldsymbol{S}'$ | A diagonal matrix. |
| $t, r$ | Transmission and reflection coefficients for a bandpass filter. |
| $t', r'$ | Transmission and reflection coefficients for a bandstop filter. |
| $\Delta, \Delta'$ | Determinant of $\boldsymbol{Z}$ and $\boldsymbol{Z}'$, respectively. |
| $\Delta_{11}, \Delta_{1n}, \Delta_{nn}$ | Co-factors of $\boldsymbol{Z}$. |
| $\Delta'_{11}, \Delta'_{1n}, \Delta'_{nn}$ | Co-factors of $\boldsymbol{Z}'$. |
| $V_{n-2}$ | Determinant of the $\boldsymbol{Z}$ matrix with first, last rows and first, last columns omitted. |
| $V_n, V_{n-1}$ | $\Delta, \Delta_{11}$ with $R = 0$, respectively. |
| $\theta, \theta'$ | Arguments of $t$ and $t'$, respectively. |
| $\omega'_p, \omega'_t$ | Relative frequency at the transmission pole and zero for a bandstop filter. |

Manuscript received March 25, 1983; revised August 15, 1983.

J. -R. Qian is with the Department of Electrical Engineering, China University of Science and Technology, Hefei, Anhui, China.

W. -C. Zhuang is with Xian Institute of Radio Technology, Xian, Shanxi, China.

0018-9480/83/1200-1045$01.00 ©1983 IEEE

| | |
|---|---|
| $\omega_p, \omega_t$ | Relative frequency at the transmission pole and zero for a bandpass filter. |
| $\omega_s'$ | Relative bandwidth. |
| $\omega_d'$ | Relative frequency difference of the passband edges. |
| $\epsilon$ | Ripple constant. |
| $Q$ | External quality factor. |
| $k_{ij}$ | Coupling coefficient. |

## I. Introduction

Bandstop waveguide filters have been, almost without exception, realized by resonators, which are connected in cascade and spaced an odd multiple of a quarter wavelength apart along the waveguide [1].

These kinds of filters can be realized with Butterworth or Chebyshev polynomials. To get an elliptic function, Rhodes [2] suggested that the resonant frequencies of these cavities should be different and the spacings between them should be different and are approximately three quarter guide wavelengths. However, to the authors' knowledge, no practical construction of such a filter has been reported.

In order to improve the frequency selectivity, the number of sections of this type of filter must be increased, which leads to considerable increase in size and weight of the filters.

In this paper, a new idea for constructing bandstop waveguide filters is presented. There are only two coupling slots or a single hole by which a multiple-coupled cavity structure is coupled to the main waveguide of the filter. The multiple-coupled cavity structure has been described by several authors [3]–[7] for realizing high-performance narrow-band bandpass waveguide filters. Now, however, is the first time the same structure has been applied to realize bandstop filters.

A multiple-coupled cavity structure contains several identical cavities, and in each cavity orthogonal or dual modes are resonant at the same frequency. The couplings take place not only among the modes of different cavities, but also between the dual modes within each cavity. Its equivalent circuit is shown in Fig. 1.

The use of the dual-mode structure and the cross-coupling technique makes the present bandstop filter much smaller in size and weight.

When such a bandpass filter has the same multiple-coupled cavity structure as a bandpass filter does, and when the symmetry and identity conditions are achieved, then their transmission responses are complementary to each other. The passbands correspond to the stopbands, and the transmission poles and zeros of one filter correspond to the zeros and poles of the other.

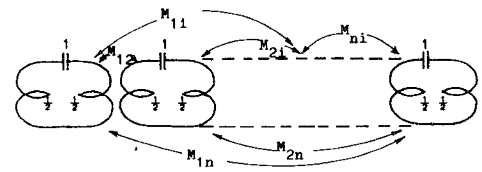

Fig. 1. Equivalent circuit for multiple-coupled cavity structure.

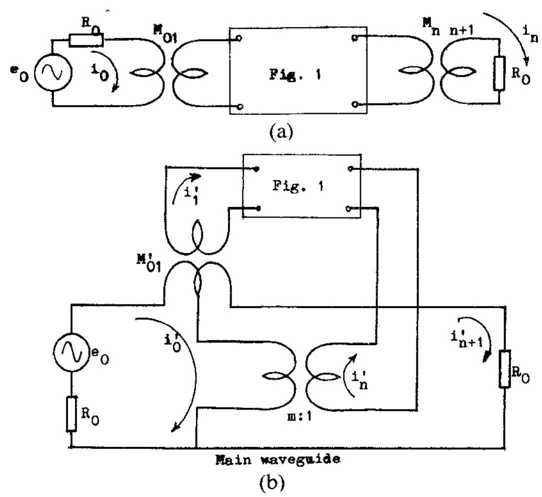


Fig. 2. Filter circuits. (a) Bandpass filter. (b) Bandstop filter.

This complementary relation makes it possible to synthesize bandstop filters by the methods of synthesizing bandpass filters with coupled cavities.

To demonstrate the principle discussed, two sample filters operating at a 5-cm wavelength were constructed. The first sample is a fourth-order antimetric elliptic bandstop waveguide filter. Its complementary filter is the bandpass filter described by Williams [3].

The second sample is a six-mode waveguide bandstop filter. The method of synthesis of the filter is the same as that described by Atia and Williams [4], [6].

## II. Theory

Multiple-coupled cavity structures which have been used to realize bandpass waveguide filters are now used to make a bandstop waveguide filter. Filter circuits, with a blank block for representing the circuit shown in Fig. 1, are illustrated in Fig. 2(a) and (b).

The equivalent circuit for the bandpass filters is shown in Fig. 2(a). The first and the last circuits in Fig. 1 are coupled to the source and the load, respectively, through the couplings $M_{01}$ and $M_{n\,n+1}$.

With reference to the equivalent circuits in Fig. 1 and Fig. 2(a), the loop equations for narrow bandwidths can be written as

$$\begin{pmatrix} e_1 \\ 0 \\ 0 \\ \vdots \\ 0 \\ 0 \end{pmatrix} = \begin{pmatrix} Z+R & jM_{12} & jM_{13} & \cdots & jM_{1n} & \\ jM_{12} & Z & jM_{23} & \cdots & JM_{2n} & \\ jM_{13} & jM_{23} & Z & \cdots & jM_{3n} & \\ \vdots & \vdots & \vdots & \cdots & Z & jM_{n-1n} \\ jM_{1n} & jM_{2n} & jM_{3n} & \cdots jM_{n-1n} & Z+R_n & \end{pmatrix} \begin{pmatrix} i_1 \\ i_2 \\ i_3 \\ \vdots \\ i_{n-1} \\ i_n \end{pmatrix} \tag{1}$$

or

$$\boldsymbol{E} = \boldsymbol{ZI}$$

where

$$R = M_{01}^2/R_0 \qquad R_n = M_{n\,n+1}^2/R_0$$

$$Z = j\left(\omega - \frac{1}{\omega}\right) \qquad e_1 = -\,je_0 M_{01}/R_0 \tag{2}$$

and the meaning of $e_0$, $R_0$ is shown in Fig. 2(a).

Further, the impedance matrix $\boldsymbol{Z}$ can be expressed as

$$\boldsymbol{Z} = \boldsymbol{S} + j\boldsymbol{M}$$

where $\boldsymbol{M}$ is an off-diagonal matrix and is termed the coupling matrix characterizing the equivalent circuit in Fig. 1. $\boldsymbol{S}$ is a diagonal matrix with elements $(Z+R, Z, Z, \cdots Z + R_n)$.

In Fig. 2(b), an equivalent circuit for the new type of bandstop filter is shown. The first and the last circuits in Fig. 1 are coupled to the main waveguide at the same place but with a 90° phase difference. Thus, if coupling $M_{01}'$ is expressed as a series mutual inductance, then coupling $M_{n\,n+1}'$ will be expressed as a shunt ideal transformer with the ratio $m:1$. This can be realized by opening a longitudinal and a transverse slot at the same place of the broadwall of a rectangular waveguide. The required phase difference is achieved because in the main waveguide the magnetic field components $H_z$ and $H_x$ (which link the fields of the dual modes through the longitudinal and the transverse slots, respectively) are 90° out of phase with each other.

The equivalent circuit for a two-slot coupler can be derived from [1, sec. 5.10]. But the equivalent circuit shown in Fig. 2(b) was derived in another paper by the authors [8]. The shunt ideal transformer results from the two cascaded impedance inverters. The one is caused by mutual inductance $M_{n\,n+1}'$ and the other can be imagined as a quarter wavelength line with characteristic impedance $R_0$, which results in a 90° phase difference between the two-slot couplers. Then it is easy to find the ratio of the ideal transformer $m = R_0/M_{n\,n+1}'$.

Referring to Fig. 1 and Fig. 2(b), the loop equations for the narrow bandwidth can be written as

$$\begin{pmatrix} e_0 \\ 0 \\ 0 \\ \vdots \\ 0 \\ 0 \end{pmatrix} = \begin{pmatrix} R_0 & jM_{01}' & 0 & 0\ldots 0 & R_0 \\ jM_{01}' & Z & jM_{12} & jM_{13}\ldots jM_{1n}\dfrac{jM_{01}'}{2m} & 0 \\ 0 & jM_{12} & Z & jM_{23}\ldots jM_{2n} & 0 \\ \cdot & \cdots & \cdots & \cdots & \cdot \\ 0 & jM_{1n}\dfrac{jM_{01}'}{2m} & jM_{2n} & jM_{3n}\ldots Z & -\dfrac{R_0}{m} \\ m & 0 & 0 & 0\ldots -1 & -m \end{pmatrix} \begin{pmatrix} i_0' \\ i_1' \\ i_2' \\ \vdots \\ i_n' \\ i_{n+1}' \end{pmatrix}. \tag{3}$$

Now let us transform the $i_0'$ loop and $i_{n+1}'$ loop, and add them to the $i_1'$ loop and the $i_n'$ loop, respectively. This can be done by using the first two and last two equations of (3), and eliminating $i_0', i_{n+1}'$ in these equations. Then we shall have

$$e_1' = (Z+R)i_1' + jM_{13}i_3' + jM_{13}i_3' + \cdots + jM_{1n}i_n'$$

$$e_n' = jM_{1n}i_1' + jM_{2n}i_2' + \cdots + (Z + R_n')i_n' \tag{4}$$

where

$$e_1' = -\,je_0 M_{01}'/(2R_0) \qquad e_n' = e_0/2m = e_0 M_{n\,n+1}'/(2R_0) \tag{5}$$

and

$$R' = M_{01}'^2/(2R_0) \qquad R_n' = R_0/2m^2 = M_{n\,n+1}'^2/(2R_0). \tag{6}$$

From (4) and the unused equations of (3), we obtain the following matrix equations:

$$\begin{pmatrix} e_1' \\ 0 \\ 0 \\ \vdots \\ e_n' \end{pmatrix} = \begin{pmatrix} Z+R' & jM_{12} & jM_{13} & \ldots jM_{1n} \\ jM_{12} & Z & jM_{23} & \ldots jM_{2n} \\ jM_{13} & jM_{23} & Z & \ldots jM_{3n} \\ \cdot & \cdot & \cdot & \cdots \\ jM_{1n} & jM_{2n} & jM_{3n} & \ldots Z + R_n' \end{pmatrix} \begin{pmatrix} i_1' \\ i_2' \\ i_3' \\ \vdots \\ i_n' \end{pmatrix} \tag{7}$$

or

$$\boldsymbol{E}' = \boldsymbol{Z}'\boldsymbol{I}' = \boldsymbol{S}'\boldsymbol{I}' + j\boldsymbol{M}\boldsymbol{I}'$$

where $\boldsymbol{M}$ is the coupling matrix and $\boldsymbol{S}'$ is a diagonal matrix with elements $(Z+R', Z, Z, \cdots Z + R_n')$.

One can readily see that the impedance matrices in (1) and (7) are similar and only the expressions for $R$, $R_n$ and $R'$, $R_n'$ are different.

From Fig. 2, the transmission and reflection coefficients for both filters can be found as follows:

$$t = (2R_0/e_0)i_{n+1} \qquad r = 1 - (2R_0/e_0)i_0. \tag{8}$$

In (8), $i_0$ and $i_n$ can be found by solving (1). When

$$i_{n+1} = jM_{n\,n+1}i_n/R_0$$

is inserted in (8), we can get the $t$ and $r$ for bandpass filters

$$t = (-1)^n \frac{2M_{01}M_{n\,n+1}}{R_0}\frac{\Delta_{1n}}{\Delta}$$

$$r = -1 + \frac{2M_{01}^2}{R_0}\frac{\Delta_{11}}{\Delta}. \tag{9}$$

When $i_0', i_{n+1}'$ resulted from (3) are inserted in (8), one readily finds that $t'$ and $r'$ for bandstop filters are

$$t' = 1 - \frac{1}{\Delta'}\left(\frac{M_{01}'}{2R_0}\Delta_{11}' + \frac{M_{n\,n+1}'}{2R_0}\Delta_{nn}'\right)$$

$$r' = \frac{1}{2R_0\Delta'}\Big(M_{01}'^2\Delta_{11}' - M_{n\,n+1}'^2\Delta_{nn}' + j(-1)^{n+1}2M_{01}'M_{n\,n+1}'\Delta_{1n}'\Big) \tag{10}$$

where $\Delta$ and $\Delta'$ are the determinants of impedance matrices in (1) and (7), respectively, $\Delta_{1n}$ and $\Delta'_{1n}$ are the co-factors of the corresponding impedance matrices with the first row and $n$th column cancelled. $\Delta_{11}$, $\Delta'_{11}$, $\Delta'_{nn}$ are also the corresponding co-factors, where the nonprime sign and prime sign apply for determinants of $\boldsymbol{Z}$ and $\boldsymbol{Z}'$ in (1) and (7), respectively.

Under the symmetry conditions

$$M_{01} = M_{n\,n+1} \qquad M_{ij} = M_{n+1-i\,n+1-j} \tag{11}$$

and identity conditions

$$\sqrt{2}\,M_{01} = M'_{01} \qquad \sqrt{2}\,M_{n\,n+1} = M'_{n\,n+1} \tag{12}$$

since $\Delta_{11} = \Delta'_{11} = \Delta'_{nn}$, $\Delta_{1n} = \Delta'_{1n}$, $\Delta = \Delta'$, one readily finds, from (9) and (10)

$$t = jr' = (-1)^n 2R\frac{\Delta_{1n}}{\Delta} \qquad r = -t' = 2R\frac{\Delta_{11}}{\Delta} - 1 \tag{13}$$

where

$$R = R_n = R' = R'_n = M_{01}^2/R_0. \tag{14}$$

As the equivalent circuit in Fig. 2 is assumed to be loss free, then from the power conservation law we have

$$|t|^2 + |t'|^2 = 1 \qquad |r|^2 + |r'|^2 = 1. \tag{15}$$

This is the complementary relation between the bandpass filter and the bandstop filter, which contain the same multiple-coupled cavity structure as expressed in Fig. 2, and for which conditions (11) and (12) are achieved.

But the complementary relations (15) are justified only in the small bandwidth, where each cavity can be treated as a single resonant circuit, and the waveguide can be considered as a transmission line.

One readily finds from (15) that the zeros and poles for the two complementary filters exchange their positions.

The relation for the time-delays of a pair of complementary filters can be derived as follows.

Let the determinant of the $\boldsymbol{Z}$ matrix with first, last rows and first, last columns omitted is $V_{n-2}$. Let determinant $V_n$ be equal to $\Delta$ with $R = 0$, and $V_{n-1}$ equal to $\Delta_{11}$ with $R = 0$. It is obvious that

$$\Delta = V_n + R^2 V_{n-2} + 2RV_{n-1} \tag{16}$$

$$\Delta_{11} = RV_{n-2} + V_{n-1}. \tag{17}$$

In the case of lossless media, $Z$ is imaginary and $M_{ij}$'s are real. So when $n$ is even, $V_n$ and $V_{n-2}$ are real, and $V_{n-1}$ and $\Delta_{1n}$ are imaginary. When $n$ is odd, the reverse is true.

If (16) and (17) are used in (13), one readily finds the arguments of $t$ and $t'$, which are expressed as

$$\theta = tg^{-1}\frac{j\left(V_n + R^2 V_{n-2}\right)}{2RV_{n-1}} \tag{18}$$

$$\theta' = tg^{-1}\frac{j2RV_{n-1}}{V_n + R^2 V_{n-2}} \tag{19}$$

respectively. From (18) and (19), it is easy to see that $tg\theta tg\theta' = -1$ or

$$\theta - \theta' = \pi/2 \pm p\pi \qquad (p = 0,1,2,\cdots). \tag{20}$$

This means that the difference of the time-delays for a pair of complementary filters is a constant.

Thus, one can conclude that two complementary filters have complementary transmission responses but same time-delay responses.

## III. Realization

In this section, we shall show how the bandstop waveguide filters can be realized by means of the principle just proved.

Of the many configurations of the multiple-coupled cavity structure, which have been used in bandpass filters, we use only the cannonical form with dual-mode folded geometries to realize the bandstop filters [5], [6]. The reasons for such a choice are that its input and output terminals are close together and this makes it possible to couple to the main waveguide at the same plane, and that the optimum transfer function (including the exact elliptic function response) can be achieved by this configuration.

To demonstrate the complementary principle, we have realized two bandstop filters with symmetrical canonical coupling set for $n = 4$ and $n = 6$, respectively.

In Williams' paper [3], a four-mode elliptic bandpass waveguide filter was described. Now we use it as a complementary one to design an elliptic bandstop waveguide filter.

Taking account of the mode order as shown in Fig. 3, we can use Williams' formulas to design such a bandstop elliptic function filter.

The first step in the design is to choose the poles $\omega'_p$ and the zeros $\omega'_z$ from given filter specifications (i.e., the relative bandwidth $\omega'_s$ and the relative frequency difference of the passband edges $\omega'_d$).

Then one must evaluate the specifications of the complementary bandpass filter. We can use (15) to find the bandpass ripple constant $\epsilon$ from the given ripple level in the stopband of the original filter. The poles $\omega_p$ and the zeros $\omega_z$ of the complementary filter are as follows:

$$\omega_p = \omega'_z \qquad \omega_z = \omega'_p.$$

The next step is to evaluate the equivalent circuit parameters (i.e., $R$, $M_{12}$, $M_{23}$, $M_{14}$) by Williams' formulas.

A $C$-band elliptic bandstop filter with relative bandwidth 2.506 percent was designed according to the procedure outlined above. From the given specifications ($\omega'_s = 0.02506$, $\omega'_d = 0.03305$), one readily finds the zero and the pole for the filter [9]

$$\omega'_z = 1.4094 \qquad \omega'_p = 0.9362.$$

Furthermore, the given ripple level in the stopband is 17 db. From these data, and according to the procedure, we find

$$R = 1.11133 \qquad M_{12} = 0.7745$$

$$M_{14} = \mp 0.4142 \qquad M_{23} = \pm 0.8603.$$

The method of realization of these parameters in waveguide cavities structure is the same as in Williams' paper if the change in mode order is taken into account.

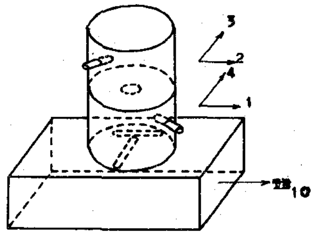

Fig. 3. A four-mode elliptic bandstop waveguide filter.

The transmission response curves are shown in Fig. 4. The theoretical and experimental results are close enough to demonstrate that the complementary principle is true for the four-mode filter. Meantime, we have realized the bandstop elliptic filter function in a waveguide structure, although its stopband attenuation is not large enough to be a practical filter. It can be improved if we start with a higher ripple level in the stopband.

The second example is a experimental six-mode bandstop filter, a photograph of which is shown in Fig. 5(a). The canonical coupling set [5] can be realized by employing dual-$TE_{111}$ circular cavity modes in three physical cavities and providing additional cross-couplings between cavities 1 and 6, and 2 and 5.

The design procedure is the same as above. However, we now use the synthesis technique described by Atia and Williams [4], [6] to design the equivalent circuit parameters. This can be done by iteratively rotating the coupling matrix and resulting in certain prescribed couplings annihilated. Then the required cavity-coupling matrix is achieved.

We start with a elliptic low-pass prototype with zeros given by $\omega_{z1} = 0.725940$ and $\omega_{z2} = 0.971499$, and poles given by $\omega_{p1} = 2.124779$, $\omega_{p2} = 1.587714$. The ripple level in the passband and the stopband for the bandstop filter are 0.05 db and 53.1 db, respectively.

Then from (15), the ripple level in the passband and the stopband for the complementary bandpass filter are 0.0000212 db and 19.4 db, respectively.

According to the same procedure described in [6], we find the parameters

$$R = 2.635461 \qquad M_{12} = 1.627265$$
$$M_{23} = 0.780838 \qquad M_{34} = -1.132448$$
$$M_{25} = 0.660570 \qquad M_{16} = -0.426915.$$

In Fig. 5(b), the cross-couplings $M_{16}$, $M_{25}$, $M_{34}$ are provided by screws, special orientations of which are arranged to provide the couplings with required signs. The $M_{23} (= M_{45})$ and $M_{12} (= M_{56})$ are realized by holes on the common walls of the cavities, the parameter $R$ (which is the loaded $Q$ of the input and output cavities) can be realized by thin transverse slot and a thin longitudinal slot, respectively.

The procedure is applied to design a bandstop filter with a bandwidth of 40 $MH_z$ at a center frequency of 4000 $MH_z$. From the relative bandwidth $\omega_s = 0.00997512$, one readily finds the external quality factor $Q$ and the coupling coefficients by

$$Q = \frac{1}{\omega_s R} \qquad k_{ij} = \omega_s M_{ij}.$$

The geometric parameters of the slots, holes, and cavities can be approximately calculated by the formulas presented in [1], [3] when $Q$'s and $k_{ij}$'s are known. But auxiliary experiments are necessary for the final determination of these parameters. The results are shown in Fig. 5(b). The theoretical and experimental transmission and return loss characteristics of the bandstop filter are plotted in Fig. 6, demonstrating the complementary principle presented, and

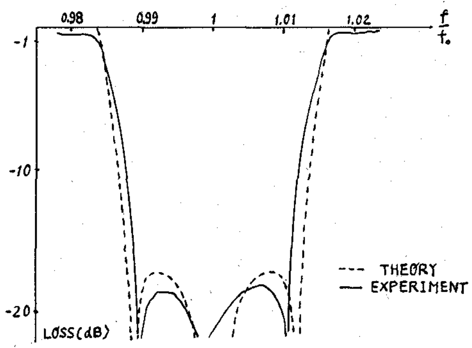


Fig. 4. Transmission characteristics of the four-mode bandstop waveguide filter.

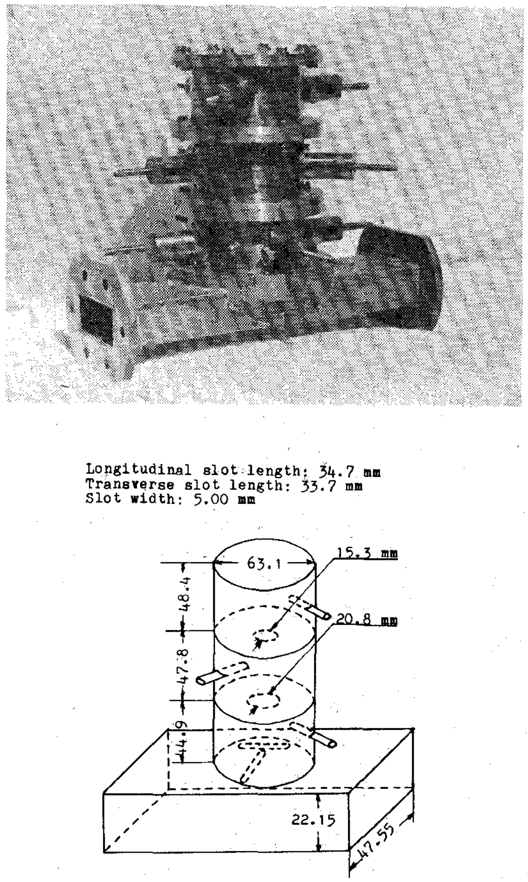


Fig. 5. A six-mode bandstop waveguide filter. (a) Photograph. (b) Construction and geometric parameters.

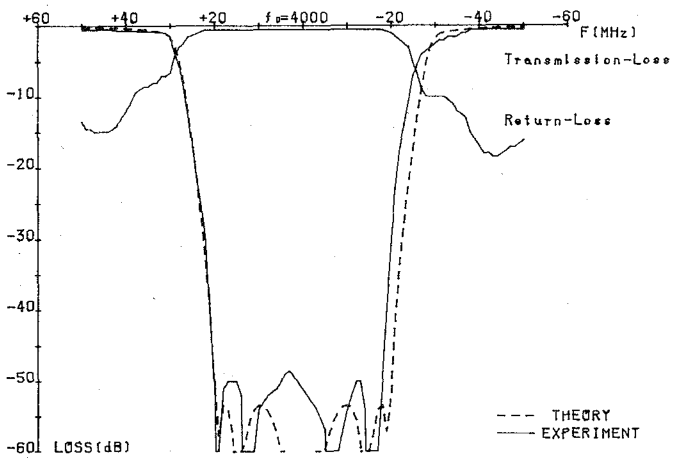


Fig. 6. Transmission and return loss characteristics of the six-mode bandstop waveguide filter.

being consistent with the expected. The experimental return loss in the passband may have been further reduced if the mismatch of the terminal could be improved.

## IV. Conclusions

A new idea for constructing a bandstop waveguide filter containing multiple-coupled cavities has been presented. The complementary principle for such a bandstop filter and a corresponding bandpass filter has been proved. By this principle, a method of synthesizing elliptic bandstop filters is described.

A four-mode and six-mode filter were constructed, and experimental results confirmed the complementary principle. Meanwhile, a practical waveguide bandstop filter was achieved.

It is also interesting to note that the complementary principle may also be applied to microstrip bandpass and bandstop filters, if a suitable coupling mechanism can be found.

## References


[1] G. L. Matthaei, L. Young, and E. M. T. Jones, *Microwave Filter, Impedance-Matching Networks and Coupling Structures*. New York: McGraw-Hill, 1964.

[2] J. D. Rhodes, "Waveguide bandstop elliptic function filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-20, pp. 715–718, Nov. 1972.

[3] A. E. Williams, "A four-cavity elliptic waveguide filter," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-18, pp. 1109–1114, Dec. 1970.

[4] A. E. Atia and A. E. Williams, "Narrow-bandpass waveguide filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-20, pp. 258–265, Apr. 1972.

[5] A. E. Williams and A. E. Atia, "Dual-mode canonical waveguide filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-25, pp. 1021–1026, Dec. 1977.

[6] A. E. Atia, A. E. Williams, and R. W. Newcomb, "Narrowband multiple coupled cavity synthesis," *IEEE Trans. Circuits Syst.*, vol. CAS-21, pp. 649–655, Sept. 1974.

[7] R. J. Cameron and J. D. Rhodes, "Asymmetric realization for dual-mode bandpass filter," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-29, pp. 51–58, Jan. 1981.

[8] W. C. Zhuang and J. R. Qian, "The duality between waveguide bandpass and bandstop filters," *CIE J. Microwaves*, no. 1, pp. 1–8, 1980.

[9] J. K. Skwirzynski, *Design Theory and Data for Electrical Filters*. London: Van Nostrand, 1965.


**Jing-ren Qian** was born in Jiangsu Province, China, on July 23, 1935. He graduated from Nanjing Institute of Technology, Nanjing, China, in 1956.

From September 1956 to May 1965, he was with the Institute of Electronics of the Chinese Academy of Sciences. From May 1965 to February 1968, he was with the Institute of Electronics of the South-West Branch of the Chinese Academy of Sciences, Chengdu. From February 1968 to January 1978, he was with Xian Radio Technology Institute, the Chinese Academy of Astronautics, where he worked on the design of microstrip circuits and waveguide filters. Currently, he is an Associate Professor in the Department of Electrical Engineering at the China University of Science and Technology, doing teaching and research work in the field of microwave devices and coupled-wave theory.

Dr. Qian is a member of the Society of Microwaves, CIE.

**Wei-chen Zhuang** was born in Shanghai, China, on June 17, 1939. He graduated from Beijing University, Beijing, China, in 1962.

After graduation, he joined the Institute of Electronics, the Chinese Academy of Sciences, where he worked on the design of microwave passive devices and did research in the field of microwave telecommunication. Since 1975, he has been with Xian Radio Technology Institute, the Chinese Academy of Astronautics, where he was involved in research concerned with satellite telecommunication and microwave multiplexers, and his current research interests include network synthesis theory and computer-aided design, as well as measurement techniques.