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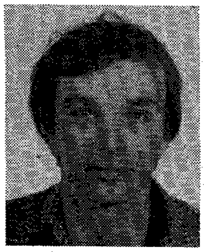
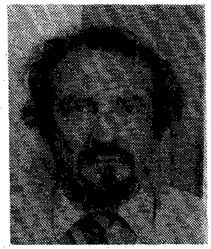


I. C. Hunter was born in Fleetwood, Lancashire, England, in 1957. From 1975–1981 he was a student at the University of Leeds, Yorkshire, England, graduating with the degrees of B.Sc. and Ph.D. in electrical and electronic engineering in 1978 and 1981, respectively. The subject of his post graduate work was electronically tunable microwave filters. During the period of his postgraduate research he was supported by the Science Research Council of the United Kingdom and by an Industrial CASE award in conjunction with Ferranti, Ltd., Dundee, Scotland.

After graduation he was employed by Filtronic Components, Ltd., Shipley, Yorkshire, England, as a Research Engineer and was involved in the development of broad-band suspended substrate filters and multiplexers. Presently, he is employed by Aercom Industries, Inc., Sunnyvale, CA, and is responsible for the development of multifunction integrated microwave modules.



John David Rhodes (M'67–F'80) was born in Doncaster, Yorkshire, England, on October 9, 1943. He received the B.Sc., Ph.D., and D.Sc. degrees in electrical engineering from the University of Leeds, Yorkshire, England, in 1964, 1966, and 1974, respectively.

He has worked as a Senior Research Engineer with Microwave Development Laboratories, Inc., Natick, MA, and has held positions of Lecturer, Reader, and Professor at Leeds University. In 1977 he formed Filtronic Components Ltd., Shipley, West Yorkshire, England, which has grown into a successful microwave company and he holds the position of Chairman, Managing and Technical Director with a part-time position of Industrial Professor at Leeds University.

Dr. Rhodes has been the recipient of five international research awards including the Browder J. Thompson, Guillimen–Cauer Awards, and the Microwave Prize. He has also been a member of several professional committees.

Analysis and Composition of a New Microwave Filter Configuration with Inhomogeneous Dielectric Medium

ATSUSHI FUKASAWA, MEMBER, IEEE

***Abstract*—A theoretical study has been undertaken to solve inter-resonator coupling problems, and to design a new structure of microwave bandpass filter.**

This filter is composed of quarter-wavelength resonators arranged in a housing with the same ends short-circuited. This arrangement of resonators has never been utilized to compose bandpass filters because of the small inter-resonator coupling. The coupling coefficient of TEM resonant lines has been derived theoretically and expressed with capacitive parameters of the coupled lines. The accurate values of capacitances of the lines were obtained by the numerical analysis of the TEM field. By a simple method to make the medium inhomogeneous through removing the dielectric partially, a simplified structure of filter has first been realized based on the theory. Through numerical analysis by the finite difference method, this structure is also shown to have higher unloaded Q of resonator than conventional dielectric-filled coaxial filters to minimize the volume of filters.

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I. Introduction

The lower regions of the microwave spectrum are recently discussed to be utilized for personal services to give communication capability at any time and at any place.

The construction of many kinds of microwave filters are based on precision metal work. The larger parts of electronic circuits in equipment are integrated and sufficiently

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low-priced by the recent semiconductor technologies. The microwave filter is now one of the parts contributing significantly to the volume, weight, and cost of radio equipment, especially of portable radio equipment.

Acoustic surface-wave technologies are not yet utilized for practical applications because of the limitations of power handling, high insertion loss, and lower selectivities.

A conventional air- or dielectric-filled coaxial resonator filter with some coupling means through windows or capacitors was first introduced to present low insertion loss and high selectivity characteristics. Any means could be found out to reduce the volume of the above filter except an improvement of high frequency loss caused by copper resistivity or medium absorption.

Recently, development of any coupling structures or studies on inter-resonator coupling problems have not been presented adequately. This paper presents a new technology giving a simple inter-resonator coupling means and a higher unloaded Q of resonators to bring out a refined structure of filter.

Theoretical approach to inter-resonator coupling problem is first discussed in this paper concerning the quarter-wavelength resonators arranged in a housing to have the same short-circuited end. This arrangement of resonators has been classified as all-bandstop response type [1]. The above arrangement of resonators was at first step tried to compose a bandpass filter under the experimental basis [2]. This paper deals with the more advanced structure under the theoretical basis. The Q of the resonator was also analyzed through the numerical calculations by the finite difference method to solve the TEM field in the structure of the coupled lines.

II. Inter-Resonator Coupling Analysis

A system of coupled transmission lines is considered to investigate the coupling problems. An inhomogeneous condition is allowed only within the cross-sectional area of the lines. Uniformity is assumed along the longitudinal direction z.

A. Propagation Constant

The schema of the system is shown in Fig. 1. The behavior of the system can be described by

$$-\frac{d}{dz}[V]=[Z][I] \tag{1}$$

$$-\frac{d}{dz}[I]=[Y][V] \tag{2}$$

where $[V]$ and $[I]$ are the two-dimensional column vectors. $[Z]$ and $[Y]$ are the matrices of impedance and admittance per unit length of the lines.

Differentiating (1) with respect to distance z and substituting into (2), a system equation of voltage on uniformly coupled lines is obtained

$$-\frac{d^2}{dz^2}[V]=[Z][Y][V]. \tag{3}$$

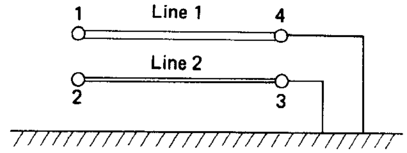
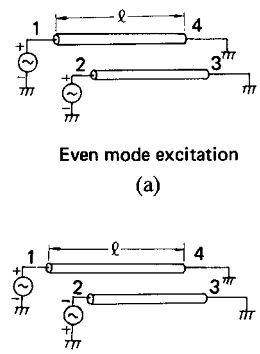


Fig. 1. Coupled transmission lines.



Even mode excitation

(a)

Odd mode excitation

(b)

Fig. 2. Eigen vectors.

The matrix product $[Z][Y]$ has the form

$$[Z][Y]=\begin{bmatrix} Z_{11}Y_{11}+Z_{12}Y_{12} & Z_{11}Y_{12}+Z_{12}Y_{22} \\ Z_{22}Y_{12}+Z_{12}Y_{11} & Z_{22}Y_{22}+Z_{12}Y_{12}\end{bmatrix}. \tag{4}$$

Solutions for (3) are obtained by solving the eigenvalue problem

$$\det|Z\cdot Y-\gamma^2E|=0 \tag{5}$$

where γ is an eigenvalue and E is the two-dimensional unit matrix.

Equation (5) leads to the following results:

$$\gamma_e^2=\left\{\frac{Z_{11}Y_{11}+Z_{22}Y_{22}}{2}+Z_{12}Y_{12}\right\} + \sqrt{(Z_{11}Y_{12}+Z_{12}Y_{22})(Z_{12}Y_{11}+Z_{22}Y_{12})} \tag{6}$$

$$\gamma_o^2=\left\{\frac{Z_{11}Y_{11}+Z_{22}Y_{22}}{2}+Z_{12}Y_{12}\right\} - \sqrt{(Z_{11}Y_{12}+Z_{12}Y_{22})(Z_{12}Y_{11}+Z_{22}Y_{12})}. \tag{7}$$

For the case $Z_{11}=Z_{22}$ and $Y_{11}=Y_{22}$, the above equations are simplified as follows:

$$\gamma_e^2=(Z_{11}+Z_{12})(Y_{11}+Y_{12}) \tag{8}$$

$$\gamma_o^2=(Z_{11}-Z_{12})(Y_{11}-Y_{12}) \tag{9}$$

where the subscripts e and o denote the even and odd excitations shown in Fig. 2. The equivalent circuit of the lines is shown in Fig. 3. L_1, L_2, C_1 and C_2 are the self inductances and capacitances of the lines 1 and 2, and L_{12} and C_{12} are the mutual inductance and capacitance between the lines 1 and 2. The following are obtained when a

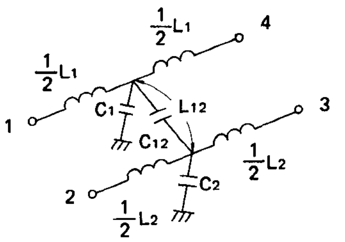


Fig. 3. Equivalent circuit of coupled transmission lines.

lossless condition is assumed:

$$[Z]=j\omega\begin{bmatrix} L_1 & L_{12} \\ L_{12} & L_2 \end{bmatrix} \tag{10}$$

$$[Y]=j\omega\begin{bmatrix} (C_1+C_{12}) & -C_{12} \\ -C_{12} & (C_2-C_{12}) \end{bmatrix}. \tag{11}$$

Substituting the above into (6) and (7), wave numbers are given as follows:

$$\beta_e=\omega\sqrt{LC}\left\{1+\frac{1}{2}\left(\frac{L_{12}}{L}-\frac{C_{12}}{C}\right)\right\} \tag{12}$$

$$\beta_o=\omega\sqrt{LC}\left\{1-\frac{1}{2}\left(\frac{L_{12}}{L}-\frac{C_{12}}{C}\right)\right\} \tag{13}$$

where $L=\sqrt{L_1L_2}$ and $\sqrt{(C_1C_{12})(C_2+C_{12})}$, respectively.

For the case $L_1=L_2$ and $C_1=C_2$, the above equations are simplified as follows from (8) and (9):

$$\beta_e=\omega\sqrt{(L_1+L_{12})C_1} \tag{14}$$

$$\beta_o=\omega\sqrt{(L_1-L_{12})(C_1+2C_{12})}. \tag{15}$$

B. Eigen Resonance Frequency

The coupled lines in Fig. 1 are supposed to be shorted at one end and opened at the other end. The specified arrangement of resonators is called here as the combline arrangement, when the open-terminal pair 1 and 2 are on one end and the short-terminal pair 3 and 4 are on the other end in a housing of filter.

Only one resonance frequency is observed experimentally for a large distance L between the lines. On the contrary, two resonance frequencies appear for a close distance L corresponding to eigen vectors shown in Fig. 2. The resonance conditions are presented as follows:

$$\beta_e l=\pi/2 \tag{16}$$

$$\beta_o l=\pi/2. \tag{17}$$

The length of the lines are supposed to be a quarter-wavelength long at each resonance frequency. Considering $L_1, L_2 \gg L_{12}$ and $C_1, C_2 \gg C_{12}$, the following are obtained:

$$\omega_e=\frac{\pi}{2l}\cdot\frac{1}{\sqrt{LC}}\left\{1-\frac{1}{2}\left(\frac{L_{12}}{L}-\frac{C_{12}}{C}\right)\right\} \tag{18}$$

$$\omega_o=\frac{\pi}{2l}\cdot\frac{1}{\sqrt{LC}}\left\{1+\frac{1}{2}\left(\frac{L_{12}}{L}-\frac{C_{12}}{C}\right)\right\}. \tag{19}$$

C. Coupling Coefficient

Equations (18) and (19) give the solution of inter-resonator coupling between resonators with the same ends short-circuited. The total inductive and capacitive coupling coefficients k_T, k_L, and k_C are given by the following:

$$k_T=k_L-k_C \tag{20}$$

$$k_L=\frac{L_{12}}{L} \tag{21}$$

$$k_C=\frac{C_{12}}{C}. \tag{22}$$

Using the above constant, the following are obtained:

$$\omega_e=\omega_c(1-k_T/2) \tag{23}$$

$$\omega_o=\omega_c(1+k_T/2) \tag{24}$$

where ω_c is the degenerated frequency between ω_e and ω_o

$$\omega_c=\frac{\pi}{2l}\cdot\frac{1}{\sqrt{LC}}. \tag{25}$$

Equations (23) to (25) lead to a known relation [3]

$$k_T=\frac{\omega_o-\omega_e}{\omega_c}=2\frac{\omega_o-\omega_e}{\omega_o+\omega_e}. \tag{26}$$

D. Capacitive Representation of Coupling Coefficient

The inductive constants of the lines can be replaced by the corresponding capacitive constants when the transverse electromagnetic field is supposed on the coupled lines.

In this paper, the term "homogeneous" is restricted to the state being fully filled with dielectric. On the contrary, "inhomogeneous" represents the state partially filled with air. The conditions $L_1=L_2$ and $C_1=C_2$ are supposed hereafter to simplify the numerical treatment. Inductances L_1 and L_{12} are found constant regardless of homogeneous or inhomogeneous conditions of the medium and they are represented by capacitive parameters C_1 and C_{12}.

For the inhomogeneous case, the following are obtained (Appendix I and II):

$$k_{L(i)}=\frac{C_{12(h)}}{C_{1(h)}+C_{12(h)}} \tag{27}$$

$$k_{C(i)}=\frac{C_{12(i)}}{C_{1(i)}}+C_{12(i)} \tag{28}$$

$$k_{T(i)}=\frac{C_{12(h)}}{C_{1(h)}+C_{12(h)}}-\frac{C_{12(i)}}{C_{1(i)}}+C_{12(i)} \tag{29}$$

where the subscripts (h) and (i) represent homogeneity and inhomogeneity.

III. Analysis by the Finite Difference Method

The coupling analysis reaches the calculation of the self and the mutual capacitances of the coupled lines. The most simplified inhomogeneous structure of filter is shown in Fig. 4 [4].

In advance of the analysis on the fields of the inhomogeneous transmission lines, a direct numerical analysis by the

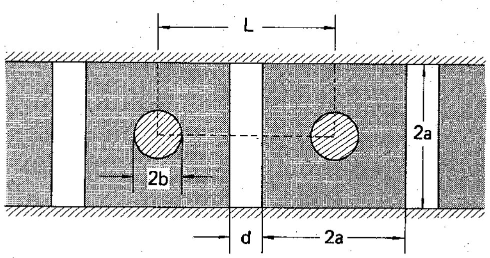


Fig. 4. Cross section of coupled lines.

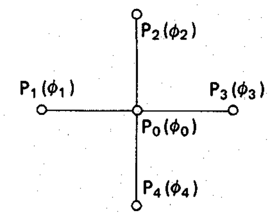


Fig. 5. Lattice points and potentials.

finite difference method was chosen for the most convenient procedure.

A. Self and Mutual Capacitances

From the Gaussian theorem in dielectric, the following is obtained:

$$q = \oint_l \epsilon_0 \epsilon_r \left(- \frac{\partial \phi}{\partial n} \right) dn \tag{30}$$

and

$$C = \frac{q}{V} = \frac{1}{V} \oint_l \epsilon_0 \epsilon_r \left(- \frac{\partial \phi}{\partial n} \right) dn \tag{31}$$

where V is the potential difference between two conductors, q and C are the surface charge of each conductor and the capacitance between two conductors per a single conductor per unit length. ϕ and n is the potential and the normal vector along a path of integration l. Two kinds of capacitances C_e and C_o correspond to a pair of orthogonal excitations shown in Fig. 2. The self and the mutual capacitances C_1 and C_{12} are obtained using (A.5) in Appendix II.

B. Unloaded Q

Attenuation constant α_c of a line caused by copper resistivity is obtained as follows:

$$\alpha_c = \frac{\frac{1}{2} Rm \frac{\epsilon_0}{\mu_0} \oint_{l_1 + l_2} \epsilon_r \left(- \frac{\partial \phi}{\partial n} \right)^2 dl}{v_i V \oint_{l_2} \epsilon_0 \epsilon_r \left(- \frac{\partial \phi}{\partial n} \right) dl} \text{(Np/m)} \tag{32}$$

where Rm is surface resistivity, μ_0 is intrinsic permittivity, v_i is the phase velocity along the line, and l_1 and l_2 are the paths of integration along the inner and the outer conductor, respectively.

The attenuation constant α_d of a line caused by dielectric absorption is also obtained as

$$\alpha_d = \frac{\frac{1}{2} \omega \epsilon_0 \epsilon_r \tan\delta \iint_s \left\{ \left(\frac{\partial \phi}{\partial n} \right)^2 + \left(\frac{\partial \phi}{\partial y} \right)^2 \right\} dS}{v_i V \oint_{l_2} \epsilon_0 \epsilon_r \left(- \frac{\partial \phi}{\partial n} \right) dl} \text{(Np/m)} \tag{33}$$

where ω is the angular frequency, $\tan\delta$ is the tangential component of dielectric loss angle, and S is the cross-sectional area of integration of the line.

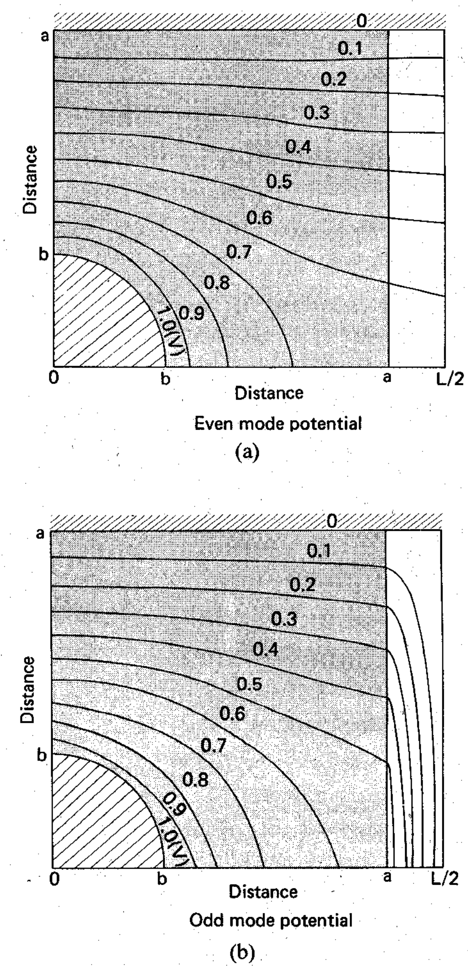


Fig. 6. Equipotential distribution of coupled lines $2a = 12$ mm, $2b = 4$ mm, $d = 4$ mm, and $\epsilon_r = 20$.

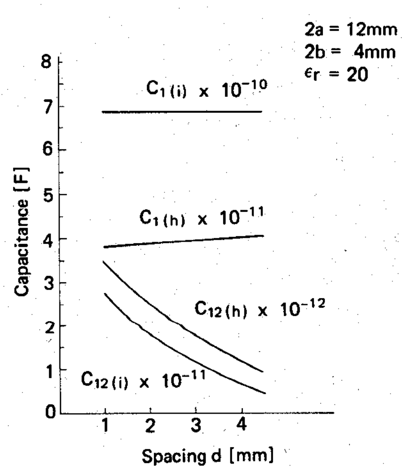
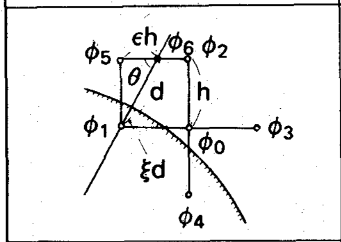
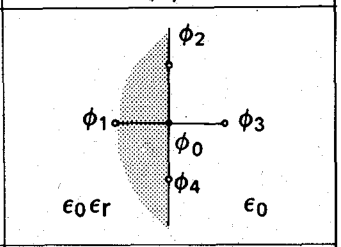


Fig. 7. Self and mutual capacitance of coupled lines.

TABLE I
PSEUDOPOTENTIAL OF THE LATTICE INCLUDING A BOUNDARY

(1)	(2)
	
$\phi'_1 = \dfrac{\phi_1 - \{\phi_5 + \epsilon(\phi_2 - \phi_5)\}}{1 - \xi}$ $\epsilon = h \cdot \cot\theta$	$\phi'_1 = \dfrac{2\epsilon_r}{\epsilon_r + 1}\phi_1$ $\phi'_3 = \dfrac{2}{\epsilon_r + 1}\phi_3$

The total attenuation constant α is

$$\alpha = \alpha_c + \alpha_d (\text{Np/m}) \tag{34}$$

and the Q factor of the line is represented as follows:

$$Q = 27.3/\alpha_\lambda, \alpha_\lambda = 8.686\alpha\lambda_g$$
$$\lambda_g = \lambda_0/\sqrt{\epsilon_{\text{eff}}} \tag{35}$$

where ϵ_{eff} is the effective dielectric constant defined by the capacitance ratio for the medium of the dielectric and the air.

C. Numerical Calculation

A cross section of inhomogeneous coupled lines is shown in Fig. 4. $2a$ is the distance between the outer conductors, $2b$ is the inner conductor diameter, d is the air spacing distance, and L is the pitch of the coupled lines. For the analysis of the field in the line, a periodic arrangement of transmission line is assumed [4]. The cross-sectional area shown in Fig. 4 is first covered by a square lattice shown in Fig. 5 and the value of each potential ϕ_0 of the point P_0 is calculated by the finite difference method [4], [5]. The size of the lattice h was determined as $h = 0.5 \sim 0.25$ (mm) and the maximum residual error R_0 was determined as $R_0 = 0.001$ (V). The value of each boundary potential was calculated using pseudopotential equations shown in Table I.

The equipotential distributions were obtained and shown in Fig. 6 as the result of calculations.

The calculated values of the self and the mutual capacitances for homogeneous and inhomogeneous conditions are shown in Fig. 7.

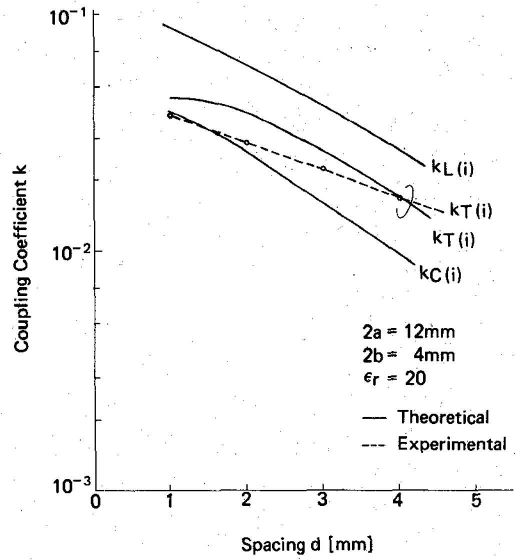


Fig. 8. Coupling coefficient of quarter-wavelength lines.

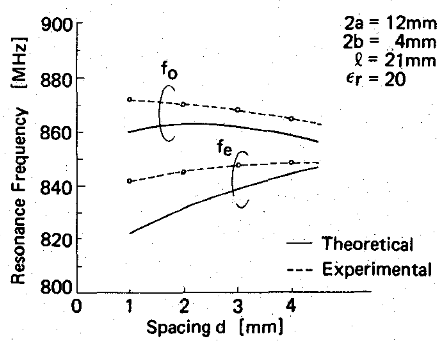


Fig. 9. Eigen resonance frequencies of quarter-wavelength lines.

IV. COMPARISON OF THEORY AND EXPERIMENT

A. Coupling Coefficient and Eigen Resonance Frequency

The coupling coefficients and a pair of eigen resonance frequencies of quarter-wavelength resonators with combline arrangement are shown in Figs. 8 and 9. The length of a resonator is 21 mm long. Referring to (8), (9), and (29), when the medium is homogeneously filled with dielectric, total coupling coefficient $k_{T(h)}$ is not zero exactly. On the contrary, when the dielectric is removed partially to be inhomogeneous, the total coupling coefficient $k_{T(i)}$ increases rapidly. Against the small value of spacing d, $k_{T(i)}$ should fall down to zero. But it was not shown in Fig. 8, because it is not easily confirmed by numerical calculations through the difficulty of conver-

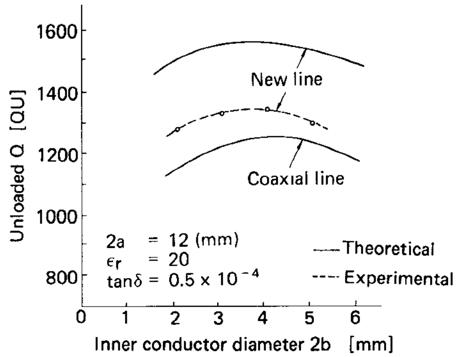


Fig. 10. Unloaded Q of coupled lines.

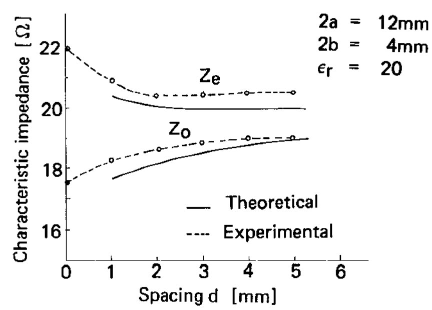


Fig. 11. Characteristic impedance of coupled lines.

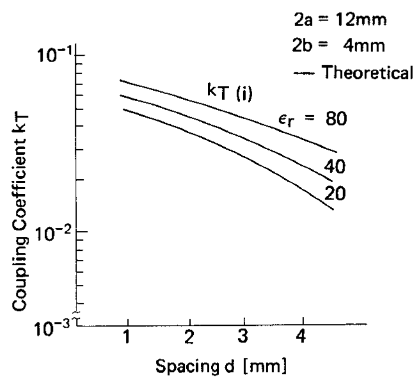


Fig. 12. Coupling coefficient versus dielectric constant.

gence of calculation. The measurement of the coupling coefficient was based on the method by M. Dishal [3].

B. Q and Characteristic Impedance

Q values of the coupled lines and a dielectric-filled coaxial line with square outer conductor are shown in Fig. 10. The measurement of Q of the line was made for a quarter-wavelength resonator neglecting the loss of the short-circuited end. For the design condition to obtain equal Q value, the area of cross section of the line is reduced to about 63 percent of that of a coaxial line. Fig. 11 shows the characteristic impedances of the lines.

C. Coupling Coefficient versus Dielectric Constant

Dielectric materials, of which dielectric constants ϵ_r ranging from 20–80, are recently available. The total coupling coefficient $k_{T(i)}$ is shown in Fig. 12 against a parameter of dielectric constant. The combline arrangement is found now capable of composing a bandpass filter having bandwidth of up to several percent by selecting value of the parameters.

V. Discussion

Equation (29) is nearly expressed by the following equation:

$$k_{T(i)} \approx \frac{C_{12(h)} - C_{12(i)}}{C_{1(h)}} \tag{36}$$

where the following are assumed: $C_{1(i)} \simeq C_{1(h)}$ and $C_{1(h)} \gg C_{12(h)}$.

The total coupling is approximately proportional to the difference value between homogeneous and inhomogeneous capacitance C_{12}.

The structure shown in Fig. 4 is understood to be one of the simple structures to produce the effect of increasing the

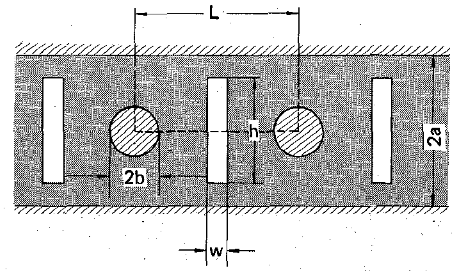


Fig. 13. Another inhomogeneous configuration.

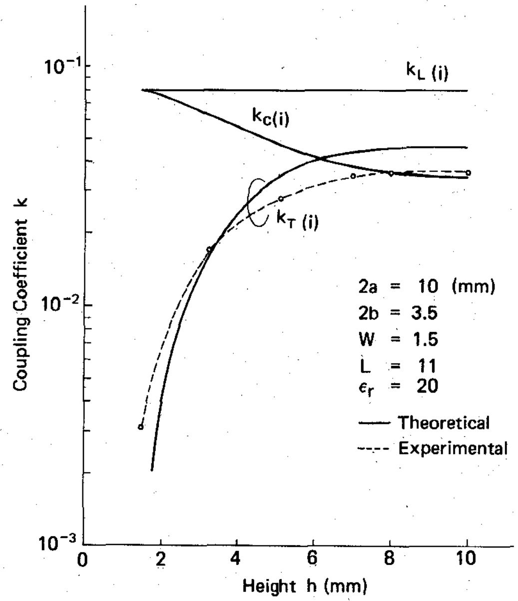


Fig. 14. Coupling coefficient of coupled lines.

coupling between resonators. Another inhomogeneous periodic cross-sectional structure is shown in Fig. 13. For the dimensions of the height h and the width w of the air gap, the coupling coefficients are determined. The theoretical and the experimental coupling coefficient are shown in Fig. 14.

VI. Design Example of Bandpass Filter

Using the results obtained relative to the basic section of coupled lines, an example of bandpass filter was designed by the following procedure.

A. Design

The synthetic design method was based on the low-pass prototype filter [6], [7].

For the expected response of bandpass filter with the attenuation α_r (dB) at the passband edge and the slope of attenuation response curve, the number of elements n of the low-pass prototype filter is uniquely fixed. The individual values g_i of the elements of the prototype low-pass filter are computed against the numerous n and α_r. The number n is assumed odd, and α_r is defined as the ripple level of the Chebyshev response. The inter-resonator coupling coefficient k, and the external Q at the input and the output terminals are used as the design parameters of bandpass filter

$$k_{i,i+1} = w/\sqrt{g_i \cdot g_{i+1}} \quad (37)$$

$$Q\text{ext}_{0,1} = g_0 \cdot g_1 / w = Q\text{ext}_{n,n+1} \quad (n\text{: odd}) \quad (38)$$

where w is the normalized bandwidth defined by

$$w = (f_2 - f_1)/f_0 \quad (39)$$

where f_0, f_1, and f_2 are the center and the passband edge frequencies.

The following conditions are specified; for example: $n = 5$, $\alpha_r = 0.02$ dB, $f_0 = 850$, $f_1 = 834$, and $f_2 = 866$ MHz. The design values are all shown in Table II, where the number $i = 0$ and $i = n + 1$ represent the input and the output terminals and the values in the parentheses show the external Q values. In Table II, the design values of the spacing d in Fig. 4 are also shown by referring to Fig. 8.

B. Performance

The characteristics of the bandpass filter are shown in Fig. 15. The attenuation response curve is found to have good symmetry against the center frequency. This is brought

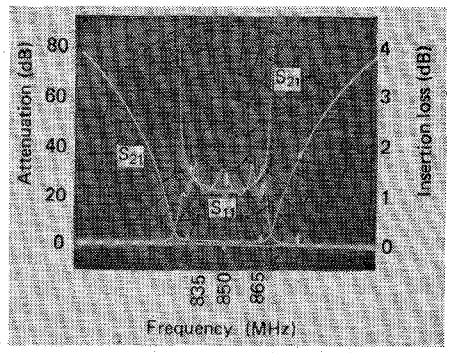


Fig. 15. Characteristics of an 800-MHz bandpass filter.

TABLE II
EXAMPLE OF DESIGN PARAMETERS OF THE BANDPASS FILTER

i	g_i	$k_{i,i+1}$	$h_{i,i+1}$ (mm) Theoretical	$h_{i,i+1}$ (mm) Experimental
0	1			
		(22.146)	—	—
1	0.886			
		0.0356	2.2	1.0
2	1.422			
		0.0248	3.2	2.6
3	1.830			
		0.0231	3.5	3.0
4	1.637			
		0.0231	3.5	3.0
5	1.830			
		0.0248	3.2	2.6
6	1.422			
		0.0356	2.2	1.0
7	0.886			
		(22.146)	—	—
8	1			

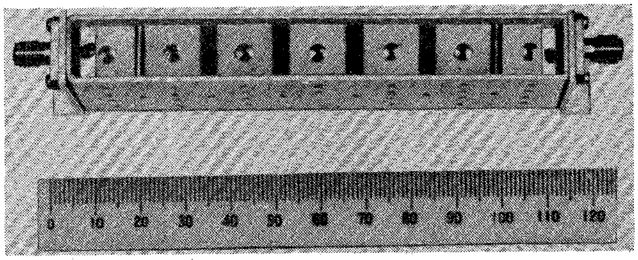


Fig. 16. 800-MHz bandpass filter.

through the distributed constant coupling effect superior to the lumped constant coupling through the capacitors, etc.

Considering the bandwidth in this example, the input and the output coupling were, however, taken using capacitive electrodes formed on the open-circuited end of the resonators. The dimensions of the electrode were determined by the experiments.

Insertion loss of a filter is about 1.2 dB and the 0.5-dB flatness bandwidth is about 32 MHz as shown in Fig. 15. The attenuation is 35 dB at $f_0 \pm 30$ MHz when f_0 is the center frequency. The volume and the weight of the filter having seven resonators are about 40 cm^3 and 170 g.

C. Feature of the New Filter

The newly developed filter shown in Fig. 16 is miniaturized by one half of the conventional dielectric-filled coaxial structure of a resonator filter.

The volume reduction was brought through higher Q of resonator and a simplified coupling structure. The higher values of Q and the coupling coefficient are basically generated by giving free conditions on the half part of the dielectric surface of a square resonator.

The tuning capacitors in conventional combline filters are not used, and shorted ends of resonators are placed on a single basement of housing. The new filter is constructed simply as compared with the conventional combline or interdigital line filter. The above features were found effective to reduce significantly the manufacturing costs of filters.

VII. Conclusion

The coupling coefficient was theoretically obtained by the eigenvalues of propagation constants through numerical analysis by the finite difference method. The theory has shown good agreement with the experiments.

Using the results of analysis, a simplified arrangement of resonance elements was found to be feasible to construct bandpass filters by placing narrow air space in the medium between resonators. The reduced space of inter-resonator coupling and a higher unloaded Q of resonator were confirmed to help reduce the volume and the cost, less than one half of that of the conventional dielectric-filled coaxial filters in 800-MHz band.

The coupling problem in interdigital line filters could not be solved by the manner presented in this paper, because the mutual position between resonators should be considered entirely under three dimensional axis (cross section and longitude).

Appendix I

Proof of $L_{(h)} = L_{(i)}$

The following relations are considered on a TEM transmission line:

$$v_{(i)} = \frac{1}{\sqrt{L_{(i)}C_{(i)}}} \quad v_{(h)} = \frac{1}{\sqrt{L_{(h)}C_{(h)}}} \tag{A.1}$$

$$\frac{\epsilon_{\text{eff}}}{\epsilon_r} = \frac{C_{(i)}}{C_{(h)}} = \left(\frac{v_{(h)}}{v_{(i)}}\right)^2. \tag{A.2}$$

Substituting (A.2) into (A.1), the following is obtained:

$$L_{(i)} = 1/C_{(i)}v_{(i)}^2 = 1/C_{(h)}v_{(h)}^2 = L_{(h)}. \tag{A.3}$$

Appendix II

Capacitive Representation of L_{12}/L_1

The following relation also exists on a TEM transmission line:

$$L_{(h)}C_{(h)} = \epsilon\mu_0. \tag{A.4}$$

For (14) and (15), the following are obtained:

$$\begin{aligned} C_{(h)e} &= C_{1(h)} \\ C_{(h)o} &= C_{1(h)} + 2C_{12(h)} \\ L_{(h)e} &= L_{1(h)} + L_{12(h)} \\ L_{(h)o} &= L_{1(h)} - L_{12(h)}. \end{aligned} \tag{A.5}$$

$$\begin{aligned} L_{1(h)} &= \frac{L_{(h)e} + L_{(h)o}}{2} = \frac{\epsilon\mu_0}{2}\left(\frac{1}{C_{(h)e}} + \frac{1}{C_{(h)o}}\right) \\ &= \frac{\epsilon\mu_0}{2}\left(\frac{1}{C_{1(h)}} + \frac{1}{C_{1(h)} + 2C_{12(h)}}\right) \\ &= \epsilon\mu_0 \frac{C_{1(h)} + C_{12(h)}}{C_{1(h)}\left(C_{1(h)} + 2C_{12(h)}\right)}. \end{aligned} \tag{A.6}$$

$$\begin{aligned} L_{12(h)} &= \frac{L_{(h)e} - L_{(h)o}}{2} = \frac{\epsilon\mu_0}{2}\left(\frac{1}{C_{(h)e}} - \frac{1}{C_{(h)o}}\right) \\ &= \frac{\epsilon\mu_0}{2}\left(\frac{1}{C_{1(h)}} - \frac{1}{C_{1(h)} + 2C_{12(h)}}\right) \\ &= \epsilon\mu_0 \frac{C_{12(h)}}{C_{1(h)}\left(C_{1(h)} + 2C_{12(h)}\right)}. \end{aligned} \tag{A.7}$$

Considering (A.3), the following result is obtained:

$$\frac{L_{12(i)}}{L_{1(i)}} = \frac{C_{12(h)}}{C_{1(h)} + C_{12(h)}}. \tag{A.8}$$

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Synthesis and Realization of Multisection Tandem Stripline Bandpass Filters

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Abstract—**A bandpass filter is described which is made up of two identical multi-element coupled symmetrical striplines in tandem connection. For the filter, a precise synthesis procedure is presented reducing the design to the synthesis of a directional coupler. The required equal-ripple polynomials are calculated by an iterative method. Relations for the polynomial extreme values are provided, derived from the attenuation requirements in passband and stopband. On the basis of this procedure the coupling factors of two 21-element filters are calculated and realized in three-layer polyolefin. Measurements show good agreement with theoretical results.**

I. Introduction

Stepped coupled TEM wavelines are widely used as directional couplers [1]–[3]. With respect to feasibility, the element with the highest coupling factor in the center of the coupling path represents the central criterion. Particularly with the 3-dB coupler, the maximum coupling factor reaches manufacturing-excluding values, because the physical spacing between the two conductors within the available ground-plane spacing is infeasible, or because the mechanical discontinuities of such a section are so great that the directivity of the coupler is severely degraded due to the interface mismatch. In such instances, high coupling elements may be overcome by using a so-called tandem interconnection of two couplers. It has been demonstrated that for most applications two tandemly connected 8.34-dB couplers are sufficient for the realization of a 3-dB coupler [3].

The problems described are exactly those of the filters consisting of stepped coupled lines. In [4], such filters are derived from directional couplers, and theoretical examples are presented. However, a practical realization of the maximum coupling factors mentioned in [4] is scarcely achievable by stripline technique. In order to arrive at a less critical manufacturing of this filter type, the following deals with such a filter in tandem connection.

The filter under consideration is schematically represented in Fig. 1. It consists of a tandem connection of two

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