Electronically Tunable Microwave Bandstop Filters

I. C. HUNTER AND JOHN DAVID RHODES, FELLOW, IEEE

Abstract —**The design procedures for varactor tuned bandstop filters are presented. A novel circuit technique for realizing bandstop filters with a symmetrical frequency response is described and explicit design formulas are presented. The physical design and experimental performance of a three cavity varactor tuned bandstop filter tunable around 4 GHz is presented. Experimental results are substantiated with computer analysis which includes the effects of varactor resistance.**

I. INTRODUCTION

THIS PAPER discusses the design of tunable bandstop filters with particular application to varactor tuning. A companion paper [1] discusses developments in the area of varactor tuned bandpass filters.

The initial design philosophy was to produce a microwave integrated circuit (MIC) varactor tuned filter. Such an approach enables ease of integration of varactor diodes and their bias chokes with the microwave filter. The particular MIC structure chosen was suspended substrate stripline (SSS), this having several advantages over microstrip and conventional stripline, e.g., higher Q factor and fewer problems with waveguide modes [2].

The SSS circuit realization of the bandstop filter is shown in Fig. 1. This consists of a uniform impedance through transmission line with capacitively coupled stubs located at intervals along it. This filter achieves high passband return loss due to the lack of discontinuities in the through line and a wide variation of stopband bandwidths can be achieved by varying the capacitive coupling gaps between the resonators and the main line. Tuning of the filter's center frequency can be achieved either mechanically or electronically by terminating the ends of the resonators in tuning screws or varactor diodes.

Of particular importance in the design of tunable bandstop filters is the effect of tuning on the stopband performance. It is necessary to examine the bandstop filter from a theoretical viewpoint in order to understand these effects. The equivalent circuit of the filter is shown in Fig. 2 where the coupling gaps are represented by pure lumped capacitances and the resonators are represented by commensurate unit elements of transmission line terminated in lumped capacitors. The main line is subdivided into unit elements of transmission line with impedance normalized to 1 Ω. It can easily be shown that the bandstop resonators possess transmission poles at frequencies close to their transmission zeros. These poles have the effect of making the frequency response of individual resonators skewed or asymmetric. Thus, when the resonators are coupled by 90° impedance inverting lengths of line the overall filter response will be asymmetric. This problem is compounded by the fact that tuning the center frequency of the filter will change the effective phase shift between the resonators, thus producing an even more pronounced asymmetric response. The best tunable performance that can be achieved is one in which the frequency response of the filter is symmetrical at the exact center of the tuning range;

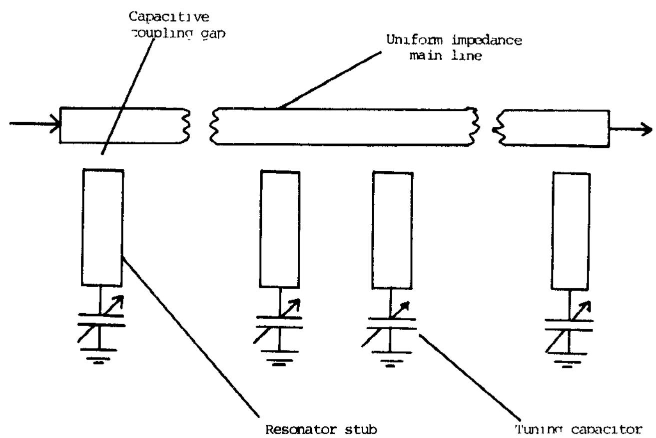

Fig. 1. Suspended stripline bandstop filter.

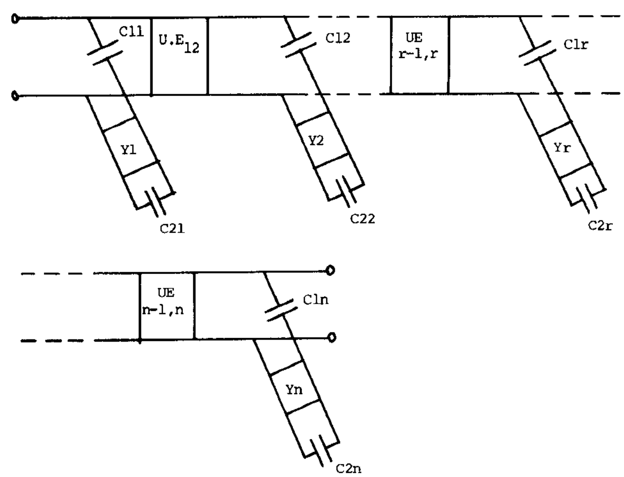

Fig. 2. The equivalent circuit of the SSS bandstop filter.

Manuscript received January 20, 1982; revised March 26, 1982.
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thus, tuning up or down in frequency will skew the response, making it more selective on the high frequency or low frequency side of the stopband, respectively. A theoretical technique to achieve this aim by evaluating the correct phase shift between the resonators is presented in Section II.

This new theoretical design technique is applied in Section III to the design of a three cavity varactor tuned bandstop filter. The techniques of realizing the filter in SSS are presented. The experimental performance of this filter was in close agreement with theoretical expectations. Also in Section III, a computer analysis of varactor tuned filters is presented; the analysis shows the effect of varactor resistance on the selectivity of the filter and enables easy determination of filter performance.

II. Theory of Symmetrical Bandstop Filters

It can easily be shown that the rth resonator shown in Fig. 2 has the following input admittance:

$$Y_r(j\omega)=\frac{j\omega C_{1r}(Z_r w C_{2r}+\tan a\omega)}{\omega Z_r(C_{1r}+C_{2r})-\tan(a\omega)(Z_r^2\omega^2 C_{1r}C_{2r}-1)}. \tag{1}$$

$Y_r(j\omega)$ possesses a pole at ω_{pr} corresponding to the rth transmission zero of the filter, and also possesses a zero at ω_{zr} where ω_{pr} and ω_{zr} are the zeros of the denominator and numerator of (1), respectively. It is the zero which occurs slightly higher in frequency than ω_{pr} which causes the asymmetric frequency response of the resonator due to the more rapid variation of $Y_r(j\omega)$ above ω_{pr} than below ω_{pr}.

For a narrow bandwidth around ω_{pr} we can represent $Y_r(j\omega)$ as follows: w

$$Y_r(j\omega)=\frac{j(\omega-\omega_{zr})}{K_r(\omega-\omega_{pr})(\omega_{zr}-\omega_{pr})} \tag{2}$$

where

$$K_r=\frac{\delta}{\delta\omega}\left|\left[Y_r(j\omega)^{-1}\right]\right|_{\omega=\omega_{pr}} \tag{3}$$

or

$$K_r=\frac{a\left[1+\tan^2(a\omega_{pr})\right]\left[Z_r^2C_{1r}C_{2r}\omega_{pr}^2-1\right]+2\tan(\omega_{pr})Z_r^2C_{1r}C_{2r}-Z_r(C_{1r}+C_{2r})}{C_{1r}\omega_{pr}(Z_r\omega_{pr}C_{2r}+\tan(a\omega_{pr}))} \tag{4}$$

and

$$\omega_{zr}\tan(a\omega_{Zr})=Z_rC_{2r} \tag{5}$$

$$\tan(a\omega_{pr})(Z_r^2\omega_{pr}^2C_{1r}C_{2r}-1)=\omega_{pr}Z_r(C_{1r}+C_{2r}). \tag{6}$$

Decomposing (2) by extracting the residue at ω_{pr} we obtain

$$Y_r(j\omega)=jB_r-j(\omega L_r-X_r)^{-1} \tag{7}$$

where

$$B_r=K_r(\omega_{zr}-\omega_{pr})^{-1} \tag{8}$$

$$L_r=K_r \tag{9}$$

$$X_r=K_r\omega_{pr}. \tag{10}$$

$Y_r(j\omega)$ has thus been approximated by a simple resonator in parallel with a frequency invariant reactance (Fig. 3). The fundamental problem is thus to remove the shunt reactance from the resonator leaving a simple resonator which will have a symmetrical frequency response. To achieve this end the admittance of each resonator is scaled (the rth resonator Y_r becoming $n_r^2Y_r$). The resonators can be represented as in Fig. 4 with

$$Y_r^1(j\omega)=-jn_r^2/(\omega L_r-X_r) \tag{11}$$

and

$$B_r=B_r'+B_r''. \tag{12}$$

Applying this procedure to every resonator and coupling the rth and $r+1$th resonators by unity impedance phase shifters of phase shift $\theta_{r,r+1}$, we obtain the typical coupling network shown in Fig. 5 with the transfer matrix

$$\begin{vmatrix}1/_{nr} & 0\\ 0 & n_r\end{vmatrix}\begin{vmatrix}1 & 0\\ jB_r'' & 1\end{vmatrix}\begin{vmatrix}\cos(\theta_{r,r+1}) & j\sin(\theta_{r,r+1})\\ j\sin(\theta_{r,r+1}) & \cos(\theta_{r,r+1})\end{vmatrix}\begin{vmatrix}1 & 0\\ jB_{r+1}' & 1\end{vmatrix}\begin{vmatrix}n_{r+1} & 0\\ 0 & 1/n_{r+1}\end{vmatrix} \tag{13}$$

$$=\begin{bmatrix}\frac{n_{r+1}}{n_r}\left[\cos(\theta_{r,r+1})-B_{r+1}'\sin(\theta_{r,r+1})\right] & \frac{j\sin(\theta_{r,r+1})}{n_r,n_{r+1}}\\ jn_rn_{r+1}\begin{bmatrix}B_r''[\cos(\theta_{r,r+1})-B_{r+1}'\sin(\theta_{r,r+1})]\\ +\sin(\theta_{r,r+1})+B_{r+1}'\cos(\theta_{r,r+1})\end{bmatrix} & \frac{n_r}{n_{r+1}}\begin{bmatrix}\cos(\theta_{r,r+1})\\ -B_r''\sin(\theta_{r,r+1})\end{bmatrix}\end{bmatrix} \tag{14}$$

which can be equated to an admittance inverter of admittance $K_{r,r+1}$ under the condition that

$$B_r''=B_{r+1}'=\cot(\theta_{r,r+1}) \tag{15}$$

with

$$K_{r,r+1}=n_r\cdot n_{r+1}/\sin(\theta_{r,r+1}) \tag{16}$$

and with initial conditions

$$n_1=1\quad B_1'=0. \tag{17}$$

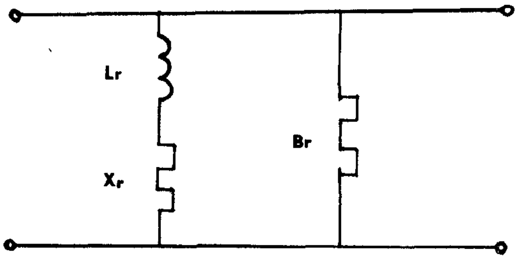

Fig. 3. The equivalent narrow-band circuit of the bandstop resonator.

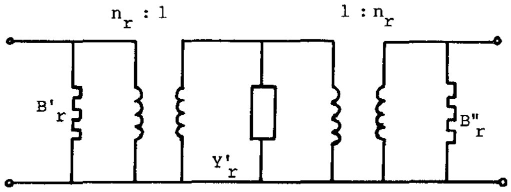

Fig. 4. Decomposition of the basic bandstop resonator.

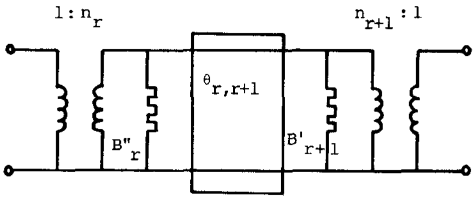

Fig. 5. The typical coupling network between the resonators of the bandstop filter.

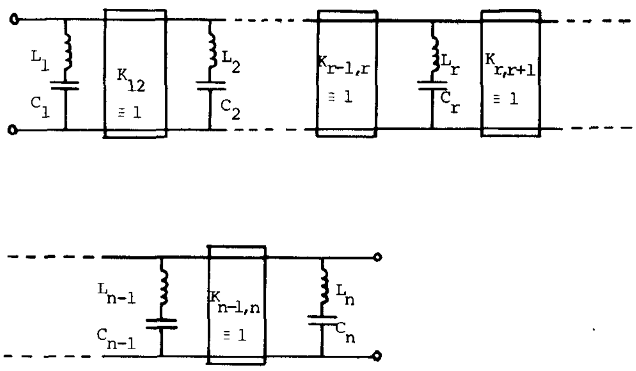

Fig. 6. The inverse Chebyshev prototype bandstop filter.

Provided the above condition is satisfied, the elements B_r in each resonator will have been absorbed into the phase shifters which become inverters. Also, the admittance of each phase shifter is unity thus retaining the uniform impedance main through line.

The design of the symmetrical bandstop filter is performed using the inverse Chebyshev prototype bandstop filter (Fig. 6). The element values for this filter are easily obtained using the method of Rhodes [3]. The resonant frequencies and bandwiths of the prototype resonators must now be equated to those of the microwave bandstop filter. The former are easily calculated (to a good approximation) as follows:

$$\omega_{or} = 1/\sqrt{L_r C_r} \tag{18}$$

$$\Delta\omega_r = 1/2L_r. \tag{19}$$

Thus, from (8) with $\omega_{pr} = \omega_{or}$ we obtain

$$\tan(a\omega_{0r})\left[Z_r^2\omega_{0r}^2 C_{1r}C_{2r} - 1\right] = \omega_{0r} Z_r\left[C_{1r} + C_{2r}\right]. \tag{20}$$

The bandwidth of the rth microwave resonator can easily be shown to be a good approximation

$$\Delta\omega_r = \left[\frac{\delta}{\delta\omega}\left(Y_r(j\omega)^{-1}\right)\right]\Bigg|_{\omega=\omega_{0r}} \tag{21}$$

which, from (6), can be seen to be equal to K_r. Thus, to obtain the resonator element values first choose the lengths of the resonators to be equal and between 90° and 180° long at the band center of the filter. Next, the lumped tuning capacitor values are chosen to have equal prescribed values. Then (6) and (22) can be solved numerically for Z_r and C_{1r}. Next, the phase shifts between the resonators at the stopband center frequency of the filter are calculated. First calculate

$$K_r = n_r^2/\Delta\omega_r. \tag{22}$$

Then evaluate ω_{zr} from (5). Next, calculate B_r from

$$B_r = \frac{1}{K_r(\omega_{zr} - \omega_{pr})} \tag{23}$$

with

$$B_r = B_r' + B_r'' \tag{24}$$

and

$$B_r'' = B_{r+1}'. \tag{25}$$

The phase shift $\theta_{r,r+1}$ can then be evaluated from

$$\theta_{r,r+1} = \cot^{-1}(B_r''). \tag{26}$$

This is a recursive procedure starting with initial conditions $n_1 = 1$ and $B_1' = 0$. After the first cycle we evaluate the new scaling factor n_2 from

$$n_{r+1} = \sin(\theta_{r,r+1})/n_r. \tag{27}$$

Since the inverse Chebyshev filter is symmetrical, this procedure need only be applied to the middle of the filter, i.e., $\theta_{12} = \theta_{n-1,n}$.

Note also that the scaling procedure results in modified resonators with $Y_r(j\omega)$ becoming $n_r^2 Y_r(j\omega)$. Thus, to retain the correct filter bandwidth the element values C_{1r}, C_{2r}, and $1/Z_r$ must be scaled by $1/n_r^2$ as a final step in the design procedure.

III. Design of Varactor Tuned Bandstop Filters

In this section, the detailed design of a varactor tuned bandstop filter using the procedure outlined in Section II and realized in SSS is presented.

A. *Choice of Varactor Diode*

The varactor diode chosen for microwave tunable filter applications must possess low resistance and a large terminal capacitance ratio. Also, the maximum capacitance of the diode must be low, of the order of one picofarad in order for realizable values of the inductive elements of the filter to be possible. The particular varactor used in this example had a zero bias junction capacitance of 0.8 pF and a capacitance ratio from zero bias to 30 V of 7.5:1. The varactor was encapsulated in a stripline package with 0.2-pF capacitance. The equivalent circuit of the diode is shown in Fig. 7, microwave resonance techniques were used to evaluate the varactor resistance which was a maximum of 1.5 Ω at zero bias and the package bondwire inductance of 500 pH.

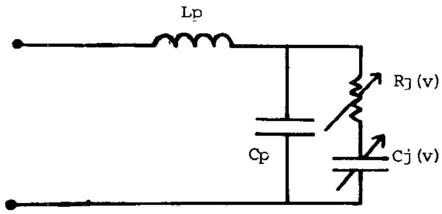

Fig. 7. Varactor equivalent circuit.

It is important to note the limitations imposed by the varactor on the tunable bandstop filter. Firstly, there is a limitation on tuning bandwidth; the package capacitance of the varactor limited the terminal capacitance ratio to 3.25:1. With the type of microwave resonator described in this paper an infinite capacitance ratio is required to produce octave tuning. With this particular varactor and a resonator phase length of 135° at the center of the tuning band, a tuning bandwidth of approximately 25 percent of the center frequency is possible. Secondly, the varactor bondwire inductance limits the maximum frequency of operation of the filter. In this particular example the bondwire will resonate the varactor at 7.1 GHz and it becomes difficult to construct filters with minimum stopband frequencies of greater than 6 GHz. Finally, there is the effect of the varactor resistance. The particular varactor described in this section had a maximum series resistance of 1.5 Ω corresponding to a Q factor of 106 at 1 GHz. This Q factor decreases linearly with increasing frequency and there are obviously limits on the selectivity of varactor tuned filters operating at microwave frequencies. For this reason computer analysis of varactor tuned filters incorporating the varactor resistance is presented. Fig. 8(a)-(c) shows the computed selectivities of the varactor tuned bandstop filter for stopband center frequencies of 1 GHz, 2 GHz, and 4 GHz as a function of degree and stopband bandwidth of the filter.

B. *Physical Design of an SSS Varactor Tuned Bandstop Filter*

A three cavity varactor tuned filter has been designed and constructed in SSS. The filter was designed for a stopband center frequency of 4 GHz and a 20-dB stopband bandwidth of 40 MHz.

The element values of the resonators and the phase shifts between them were computed using the method outlined in Section II. In this design, the resonators were chosen to be 135° long at 4 GHz and the varactor capacitance was chosen to be the geometric mean of the extremes of the terminal varactor capacitance, i.e., 0.55 pF. The phase shifts between the resonators were computed to be 82.8° at 4 GHz.

Calculation of the physical widths of the resonators and main line was performed using Getsingers technique [4]. Calculation of the coupling gaps between the resonators and the main line required to realize the lumped capacitors C_{1r} was performed using experimental techniques. The capacitance of these gaps normalized to ground plane

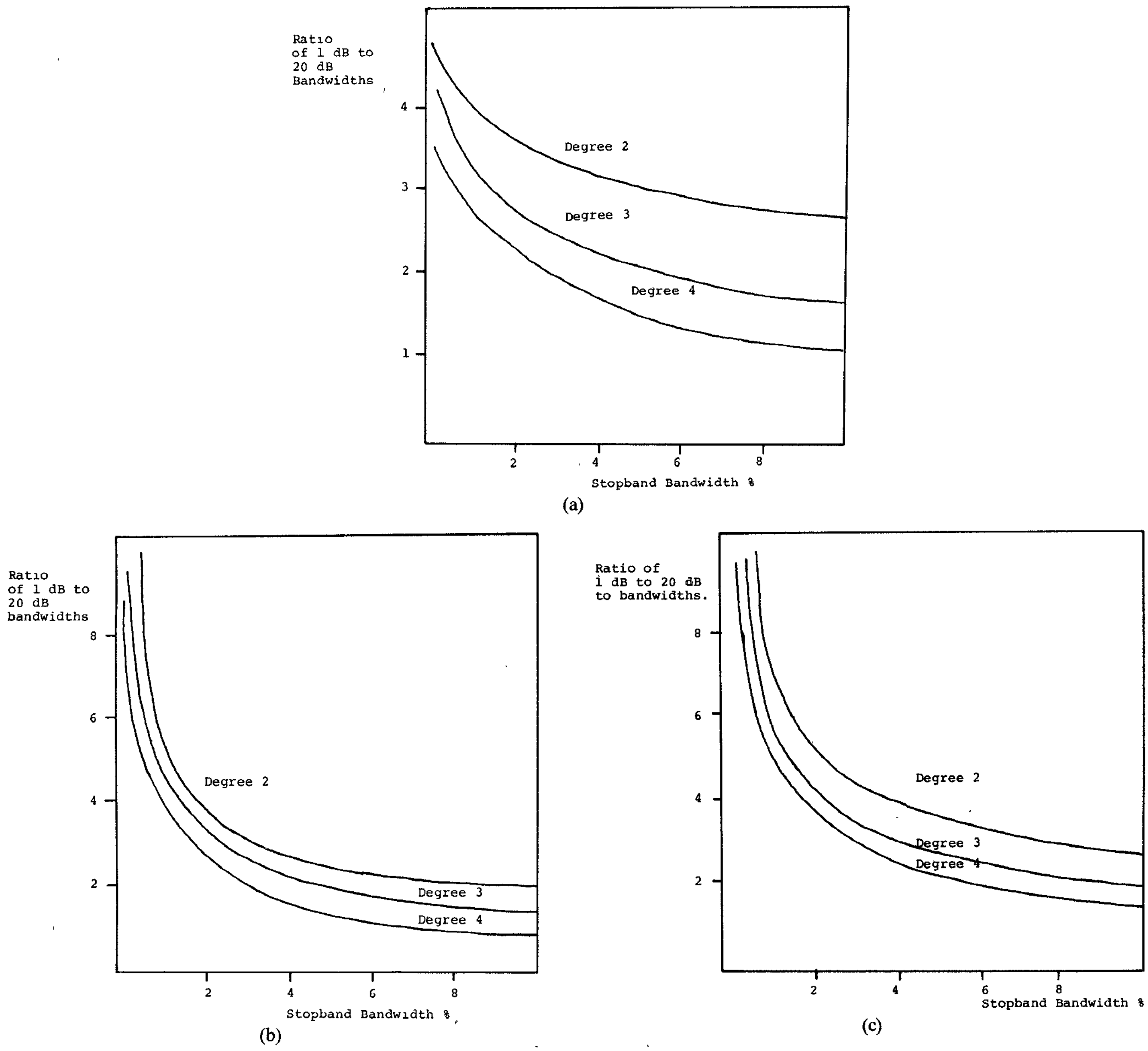

Fig. 8. (a) Ratio of 1–20-dB stopband bandwidths of a varactor tuned filter of center frequency 1 GHz as a function of stopband bandwidth and the degree of the filter. (b) Ratio of 1–20-dB stopband bandwidths of a varactor tuned filter of center frequency 2 GHz as a function of stopband bandwidth and the degree of the filter. (c) Ratio of 1–30-dB stopband bandwidths of a varactor tuned filter of center frequency 4 GHz as a function of stopband bandwidth and the degree of the filter.

spacing is shown as a function of gap spacing and resonator thickness in Fig. 9.

The symmetry of the frequency response of the bandstop filter has been shown to be critically dependent on the phase shift between the resonators. For this reason, the reference plane locations for the unit elements between the resonators must be allowed for in the design. These reference planes are shown in Fig. 10.

The bondwire inductance of the varactor diode was allowed for by modifying the lengths of the resonators by an appropriate amount.

Varactor bias chokes were realized using 150-Ω quarter-wave lines shunted at one end by 10-pF chip capacitors. These chokes perform extremely well.

The SSS circuit board and housing were gold plated to assist in the varactor ground connection. A picture of the interior of the device is shown in Fig. 11.

C. Measured Performance of the Varactor Tuned Bandstop Filter

The frequency response of the varactor tuned bandstop filter is shown in Fig. 12(a)–(c). The filter exhibited tuning from 3.5 GHz to 4.5 GHz, with the required symmetrical response at 4 GHz. It is important to note that the stop-

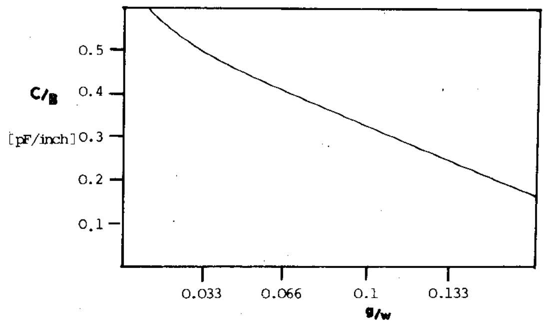

Fig. 9. The relationship between the dimensions of the resonator coupling gap and its associated lumped capacitance. The capacitance is normalized to 1-in wide resonators.

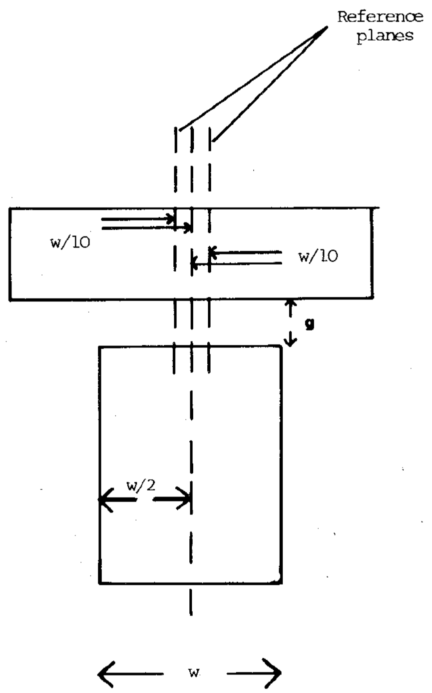

Fig. 10. Measured reference plane locations of the bandstop resonators.

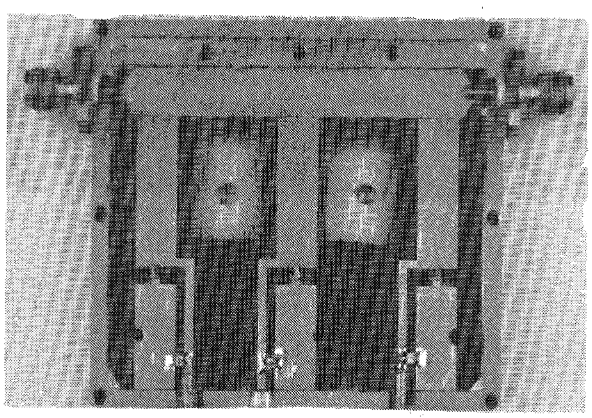

Fig. 11. Interior of filter.

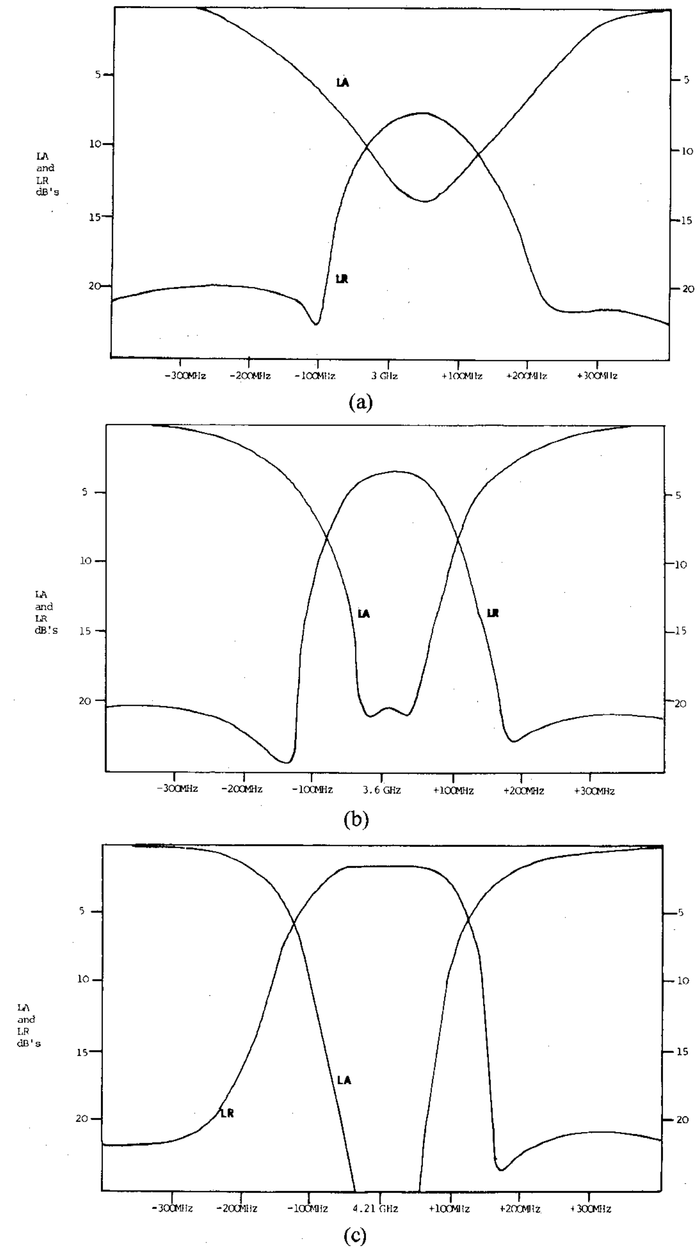

Fig. 12. (a) Filter response with OV varactor bias. (b) Filter response with 4-V bias. (c) Filter response with 30-V varactor bias.

band bandwidth of the filter increased rapidly with tuned frequency. This is a consequence of the frequency dependence of the coupling gaps between the resonators and the main line. To retain constant bandwidth it would be necessary to varactor tune the coupling capacitors.

Large signal measurements indicated that the filter had a minimum value of second-order intercept point at zero bias of +13 dBm.

IV. Conclusions

The design techniques for varactor tuned bandstop filters have been presented. A novel technique for designing bandstop filters with symmetrical frequency response has been developed and this technique has been applied to the design of a varactor tuned bandstop filter realized in SSS. This filter exhibited tuning from 3.5 GHz to 4.5 GHz and performed according to theoretical expectations. Computer analysis of varactor tuned filters showing the effect of varactor loss on filter performance was presented.

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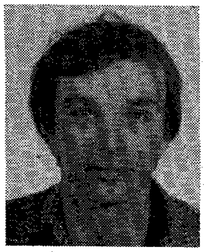

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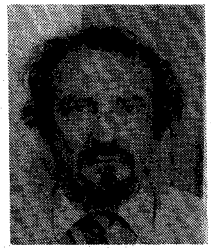

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Dr. Rhodes has been the recipient of five international research awards including the Browder J. Thompson, Guillimen–Cauer Awards, and the Microwave Prize. He has also been a member of several professional committees.

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

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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.

Manuscript received November 25, 1981; revised March 26, 1982.

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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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