Electronically Tunable Microwave Bandpass Filters

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

Abstract —**Combline filters with novel input and output coupling networks which enable broad-band tuning to be achieved with minimum degradation in passband performance are discussed. Explicit design formulas for these filters are presented. Computer analysis of varactor tuned combline bandpass filters including the small signal varactor equivalent circuit is presented enabling filter performance to be easily evaluated. The design and experimental performance of a varactor tuned combline filter, realized in suspended substrate stripline is described. This filter tuned from 3.2 GHz to 4.9 GHz exhibited low passband insertion loss and its performance was in close agreement with theoretical expectations.**

I. Introduction

ELECTRONICALLY tunable microwave filters have wide application in ESM receiving systems [1]. System requirements necessitate that the filters exhibit broad tuning bandwidths and high Q factors. These requirements are admirably met by YIG filters [2] which utilize the phenomenom of the variation in the ferromagnetic resonant frequency of Yittrium–Iron–Garnet spheres as a function of an externally applied dc magnetic field. YIG filters exhibit multioctave tuning bandwidths and resonator Q factors of up to 10 000. However, the tuning speed of YIG filters is severely limited by magnetic hysteresis effects.

Modern system developments present a need for tuning speeds in excess of 1 GHz/μs. To achieve this requirement, varactor tuning techniques must be employed. Varactor diodes utilize the change in depletion layer capacitance of a p-n junction as a function of applied bias voltage [3]. These devices do not exhibit hysteresis, and the tuning speeds of varactor tuned filters are limited only by the time constant of the varactor bias filter.

Varactor Q factors are considerably lower than those of YIG devices and thus varactor tuned filters are not generally used above 2 GHz. However, with the advent of Gallium Arsenide (GaAs) technology, varactor Q factors are steadily increasing and higher frequency operation is now feasible. This paper describes recent developments in the design of varactor tuned bandpass filters for frequencies up to 10 GHz, a companion paper [4] describes similar developments with bandstop filters.

Narrow-band microwave bandpass filters rely upon the electromagnetic coupling between resonators to provide the required impedance inverting circuit elements [5]. Such coupling is, by nature, frequency dependent and thus the impedance inverters only operate correctly over narrow bandwidths. Hence, tuning over broad bandwidths produces a deterioration in passband return loss. This paper describes a tunable combline filter in which the internal coupling frequency dependence is compensated to a large degree by correct design of the terminating transformer elements. Using this filter, octave tuning can be achieved while retaining high passband return loss. In addition, this filter possesses the important property of maintaining approximately constant absolute passband bandwidth independent of tuned frequency. Explicit design formulas are presented for this filter.

Computer analysis of the varactor tuned combline filter is presented. This analysis incorporates the varactor resistance and thereby enables the filter insertion loss as a function of frequency, passband bandwidth, and degree of the filter to be determined.

Finally, the design and experimental performance of a two cavity 5 percent bandwidth varactor tuned combline filter are presented. This filter was constructed in suspended substrate stripline and exhibited tuning from 3.2 GHz to 4.9 GHz with a maximum passband insertion loss of 6 dB and a minimum passband return loss of 8 dB.

II. Theory of Tunable Combline Filters

The combline filter is composed of a commensurate multiwire transmission line with coupling constrained to be only between adjacent lines. The lines are each short circuited to ground at the same end while opposite ends are terminated in lumped capacitors (Fig. 1). At the resonant frequency of the filter, the lines are significantly less than a quarter wavelength long, thus the filter is physically compact and it possesses a broad stopband bandwidth.

Fig. 2 shows the equivalent circuit of the combline filter. In Fig. 2, the resonators are composed of distributed inductors in parallel with lumped capacitors, while the coupling between resonators is via series distributed inductors.

If the series coupling inductors are now shunted by identical elements but with opposite sign (Fig. 3) it can be seen that the transfer matrix between the rth and $r+l$th

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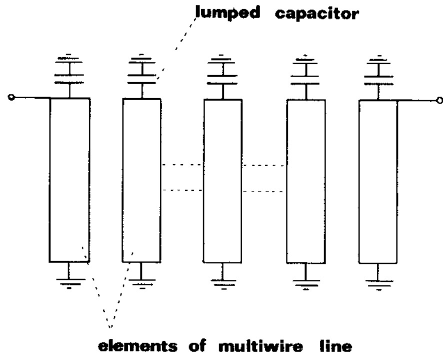

Fig. 1. The combline filter.

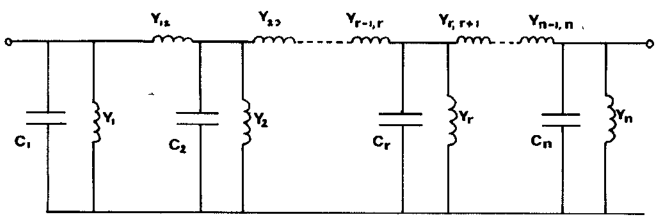

Fig. 2. Equivalent circuit of the combline filter.

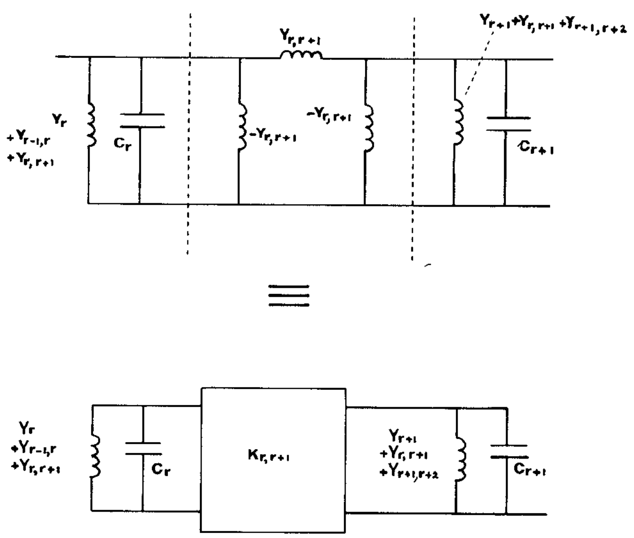

Fig. 3. Formation of impedance inverters in the combline filter.

nodes in the network will be as follows:

$$(T)=\begin{bmatrix}1 & 0\\ \dfrac{jY_{r,r+1}}{\tan(a\omega)} & 1\end{bmatrix}\begin{bmatrix}1 & \dfrac{j\tan(a\omega)}{Y_{r,r+1}}\\ 0 & 1\end{bmatrix}\begin{bmatrix}1 & 0\\ \dfrac{jY_{r,r+1}}{\tan(a\omega)} & 1\end{bmatrix}$$
$$=\begin{bmatrix}0 & \dfrac{j\tan(a\omega)}{Y_{r,r+1}}\\ \dfrac{jY_{r,r+1}}{\tan(a\omega)} & 0\end{bmatrix} \tag{1}$$

which is the transfer matrix of an admittance inverter of admittance

$$K_{r,r+1}=Y_{r,r+1}/\tan(a\omega). \tag{2}$$

The admittance of the rth resonator is

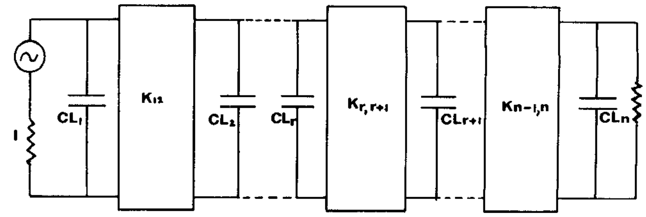

Fig. 4. The low-pass prototype filter.

$$Y_{(r)}=j(\omega C_r-(Y_r+Y_{r-1,r}+Y_{r,r+1})/\tan(a\omega)). \tag{3}$$

Now, removing the tangential frequency dependence of the inverters by scaling the entire network admittance (including the terminating resistors) by a factor $\tan(ar\omega)/\tan(a\omega 0)$ we obtain

$$K_{r,r+1}=Y_{r,r+1}/\tan(a\omega_0) \tag{4}$$

and

$$Y_{(r)}=\frac{j}{\tan(a\omega_0)}(\omega C_r\tan(a\omega)-(Y_r+Y_{r-1,r}+Y_{r,r+1})) \tag{5}$$

where ω_0 is the passband center frequency of the combline filter.

Thus from (5) and the low-pass prototype filter shown in Fig. 4, we obtain the low-pass to bandpass frequency transformation

$$\omega\rightarrow\alpha(\beta\omega\tan(a\omega)-1) \tag{6}$$

where

$$\alpha=(Y_r+Y_{r-1,r}+Y_{r,r+1})/(CL_r\tan(a\omega_0)) \tag{7}$$

and

$$\beta=C_r/(Y_r+Y_{r-1,r}+Y_{r,r+1}) \tag{8}$$
$$=1/\omega_0\tan(a\omega_0). \tag{9}$$

Equating $\omega=\pm 1$ in the low-pass prototype to the passband edges in the combline filter we obtain

$$-1=\alpha(\beta\omega_1\tan(a\omega_1)-1) \tag{10}$$
$$+1=\alpha(\beta\omega_2\tan(a\omega_2)-1). \tag{11}$$

For narrow passband bandwidths denoted $\Delta\omega$, the band-edges are approximated as follows:

$$\omega_1=\omega_0-\frac{\Delta\omega}{2} \tag{12}$$
$$\omega_2=\omega_0+\frac{\Delta\omega}{2}. \tag{13}$$

Substituting (12) into (10) and using the well-known trigonometric expansion of $\tan(A+B)$, then provided that $\Delta\omega\ll\omega_0$, we obtain

$$-1\simeq\alpha\left(\beta\left(\omega_0-\frac{\Delta\omega}{2}\right)\left(\frac{\tan(a\omega_0)-a\Delta\omega/2}{1+\dfrac{a\Delta\omega}{2}\tan(a\omega_0)}\right)-1\right). \tag{14}$$

Expanding (14), using the binomial theorem and ignoring second ordered terms in $\Delta\omega$, we obtain

$$-1=\alpha\left(\beta\left(\omega_0\tan(\theta_0)-\frac{\Delta\omega}{2}\left(\tan(\theta_0)+\theta_0\left(1+\tan^2(\theta_0)\right)\right)\right)-1\right) \quad (15)$$

where $\theta_0 = a\omega_0$. Similarly, from (11) and (13)

$$+1=\alpha\left(\beta\left(\omega_0\tan(\theta_0)+\frac{\Delta\omega}{2}\left(\tan(\theta_0)+\theta_0\left(1+\tan^2(\theta_0)\right)\right)\right)-1\right). \quad (16)$$

Solving (15) and (16) simultaneously for $\Delta\omega$ yields

$$\Delta\omega=\frac{2}{\alpha\beta\left(\tan(\theta_0)+\theta_0\left(1+\tan^2(\theta_0)\right)\right)}. \quad (17)$$

Thus, from (17) and (9) we obtain an expression for the bandwidth of the combline filter as a function of tuned frequency

$$\Delta\omega=\frac{2\omega_0\tan(\theta_0)}{\alpha\left(\tan(\theta_0)+\theta_0\left(1+\tan^2(\theta_0)\right)\right)}. \quad (18)$$

Replacing ω_0 by θ_0 in (18), and then by inspection, (18) can be seen to possess a turning point for a value of θ_0 between 0 and 90°. Using the standard differentiation procedure it can be shown that the turning point occurs at $\theta_0 = 52.885°$. This value of θ_0 corresponds to the frequency at which the bandwidth is maximized.

Δ is shown as a function of θ_0 in the following list:

θ_0	$\Delta\omega$
20°	0.167
30°	0.237
40°	0.288
50°	0.314
60°	0.306
70°	0.254.

It is thus evident that, with the combline resonators designed to be 52.885° long in the middle of the tuning band, octave tuning can then be obtained with less than a 20-percent deviation in passband bandwidth.

As yet, the frequency dependence of the admittance inverters has not been removed from the filter but merely scaled into the terminating resistors which now have a conductance

$$GL = G_s = \tan(a\omega_0)/\tan(a\omega). \quad (19)$$

Octave tuning around $\theta_0 = 52.885°$ will result in a 3.92:1 change in these elements and a corresponding drastic deterioration in the filter's return loss. This frequency variation can almost entirely be removed by introducing nonresonated transformer elements at the input and output of the filter (Fig. 5). It is proven in the Appendix that the

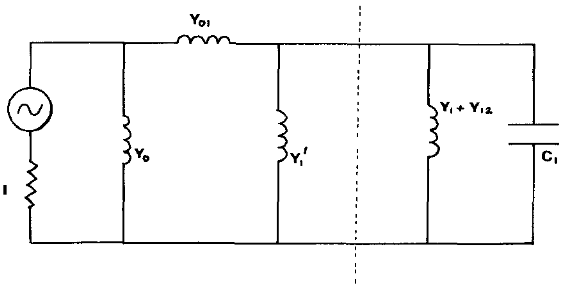

Fig. 5. Showing the transformer elements at the input and output of the combline filter.

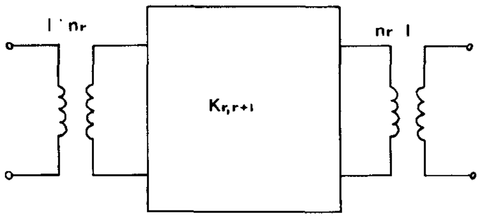

Fig. 6. The typical coupling network between internal nodes of the combline filter.

admittances of the transformer elements must satisfy

$$Y_0 = 1 - 1/\cos(\theta_0) \quad (20)$$

$$Y_{01} = 1/\cos(\theta_0) \quad (21)$$

$$Y_1 = 1 - 1/\cos(\theta_0). \quad (22)$$

Under these conditions, the load admittance looking back from the filter in YL and after scaling by $\tan(a\omega)/\tan(a\omega_0)$ is given by

$$YL=\frac{\sin(2\theta)}{\sin(2\theta_0)}+j\frac{(\cos(2\theta)-\cos(2\theta_0))}{\sin(2\theta_0)}. \quad (23)$$

YL is resonant at θ_0 and the required minimum variation in the real part of YL is achieved. Octave tuning from $\theta = 30°$ to $\theta = 60°$ will produce a maximum load variation of 0.866:1 and the deterioration in passband return loss will thus be greatly reduced. Note that the optimum value of θ_0 required to minimize the variation in YL is 45°.

The admittance level of the filter is now scaled at each internal node in order to achieve realizable element values for narrow passband bandwidths. After scaling the rth internal node by a factor nr^2 we obtain the typical coupling network between nodes r and $r+1$ shown in Fig. 6 with the following transfer matrix:

$$\begin{bmatrix}1/nr & 0\\ 0 & nr\end{bmatrix}\begin{bmatrix}0 & j/K_{r,r+1}\\ jK_{r,r+1} & 0\end{bmatrix}\begin{bmatrix}nr & 0\\ 0 & 1/nr\end{bmatrix}$$

$$=\begin{bmatrix}0 & j/_{nr\cdot nr+1}K_{r,r+1}\\ jK_{r,r+1\,nr\cdot nr+1} & \end{bmatrix}. \quad (24)$$

Thus

$$K_{r,r+1} \to {}_{nr\cdot nr+1}K_{r,r+1}$$

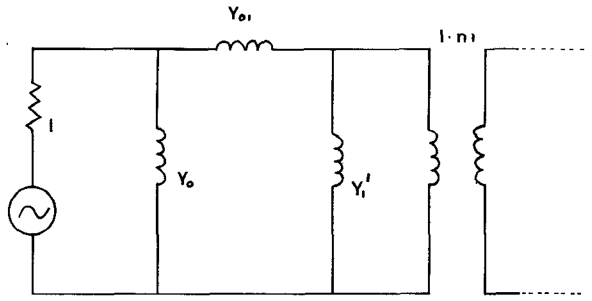

Fig. 7. Input of combline filter after scaling.

or

$$K_{r,r+1} = {}_{nr \cdot nr+1}Y_{r,r+1}/\tan(\theta_0). \tag{25}$$

The scaling constant becomes

$$\alpha = \frac{nr^2(Y_r + Y_{r-1,r} + Y_{r,r+1})}{CL_r \tan(\theta_0)}. \tag{26}$$

The ideal transformers which remain at the input and output of the filter (Fig. 7) must be absorbed into the input and output coupling. To achieve this and to preserve the correct frequency variation of YL, it can be shown by a procedure similar to that given in the Appendix that the new values of these elements must be

$$Y_0 = 1 - 1/n1\cos(\theta_0) \tag{27}$$

$$Y_{01} = 1 - 1/n1\cos(\theta_0) \tag{28}$$

and

$$Y_1 = 1/n1^2 - 1/n1\cos(\theta_0). \tag{29}$$

Note that Y_1 is negative; this is unimportant since this element can be absorbed into the first resonator of the filter.

The design equations can now be evaluated. From (8) and (9) we obtain

$$(Y_r + Y_{r-1,r} + Y_{r,r+1}) = C\omega_0 \tan(\theta_0) \tag{30}$$

where C is the capacitance of the lumped tuning capacitors at band center. These can be chosen to be equal for all resonators. From (19)

$$\alpha = \frac{2\omega_0 \tan(\theta_0)}{\Delta\omega(\theta_0 + \theta_0(1 + \tan^2(\theta_0)))}. \tag{31}$$

From (26)

$$nr = ((\beta CL_r \tan(\theta_0))/(Y_r + Y_{r-1,r} + Y_{r,r+1}))^{1/2}. \tag{32}$$

From (26)

$$Y_{r,r+1} = \frac{K_{r,r+1}\tan(\theta_0)}{nr \cdot nr + 1}. \tag{33}$$

Thus, in summary, we initially obtain the element values CL_r and $K_{r,r+1}$ for the low-pass prototype filter, normally the Chebyshev prototype [6].

Next, θ_0 must be chosen noting that for optimum return loss θ_0 should be 45° and for minimizing the bandwidth variation with tuned frequency θ_0 should be 52.885°. Next, the value of the lumped tuning capacitors at mid-tuning band should be chosen and then applying (30) to (33) for $r = 1$ to n the element values Y_r and $Y_{r,r+1}$ can be obtained. Finally the admittances of the input and output coupling elements are obtained from (27)–(29).

III. Practical Varactor Tuned Combline Filters

If packaging effects are ignored, then the reverse biased varactor diode can be approximated (for small signal levels) by the series connection of a resistor $R_j(v)$ and a capacitor $C_j(v)$ where $R_j(v)$ is the resistance of the undepleted epilayer in the device and $C_j(v)$ is the depletion region capacitance. The Q factor of this diode model is expressed by

$$Q(v) = (2\pi f C_j(v) R_j(v))^{-1}. \tag{34}$$

A high quality GaAs varactor will typically have $R_j(OV) \simeq 1.25\ \Omega$ and $Cj(OV) \simeq 1$ pF. Thus, at 1 GHz the zero-bias Q factor of such a device will be only 125 and this value decreases as the reciprocal of frequency. Thus, microwave varactor tuned filter insertion loss levels will be predominantly controlled by the varactor loss. For this reason the insertion loss of varactor tuned filters has been computed as a function of frequency, passband bandwidth, and degree of the filter. The computed results are presented graphically in Fig. 8. For the sake of brevity, results are only presented for 1-pF, 1.25-Ω, and zero-biased varactors.

A varactor tuned combline filter has been designed and constructed to the following specification:

Center Frequency	4.5 GHz ± 750 MHz
Passband Bandwidth	200 MHz ± 10 percent
Degree of Filter	2.

The varactor used in this design was a 30-V device with a zero-bias chip capacitance of 0.9 pF and a resistance of 1.3 Ω. This device, if encapsulated, would exhibit a 7:1 capacitance ratio, and from (9) we see that octave tuning is possible. However, the varactor was enclosed in a stripline package with 0.18-pF capacitance. The package capacitance reduced the terminal capacitance ratio of the filter to 3.5:1, thus the tuning bandwidth of the filter would be from $\theta = 35°$ to $\theta = 55°$, that is, from 3.5 GHz to 5.5 GHz. A varactor capacitance of 0.577 pF was chosen for the design; this value is the geometric mean of the extremes of the varactor capacitance. θ_0 was chosen to be 45°. A Chebyshev prototype filter with 15-dB passband return loss was used. This prototype had the following element values:

$$CL_1 = 0.929 = CL2$$

$$K_{12} = 1.197.$$

Now $C = 2.88 \times 10^{-11}$ (normalized to a 1-Ω system). Thus,

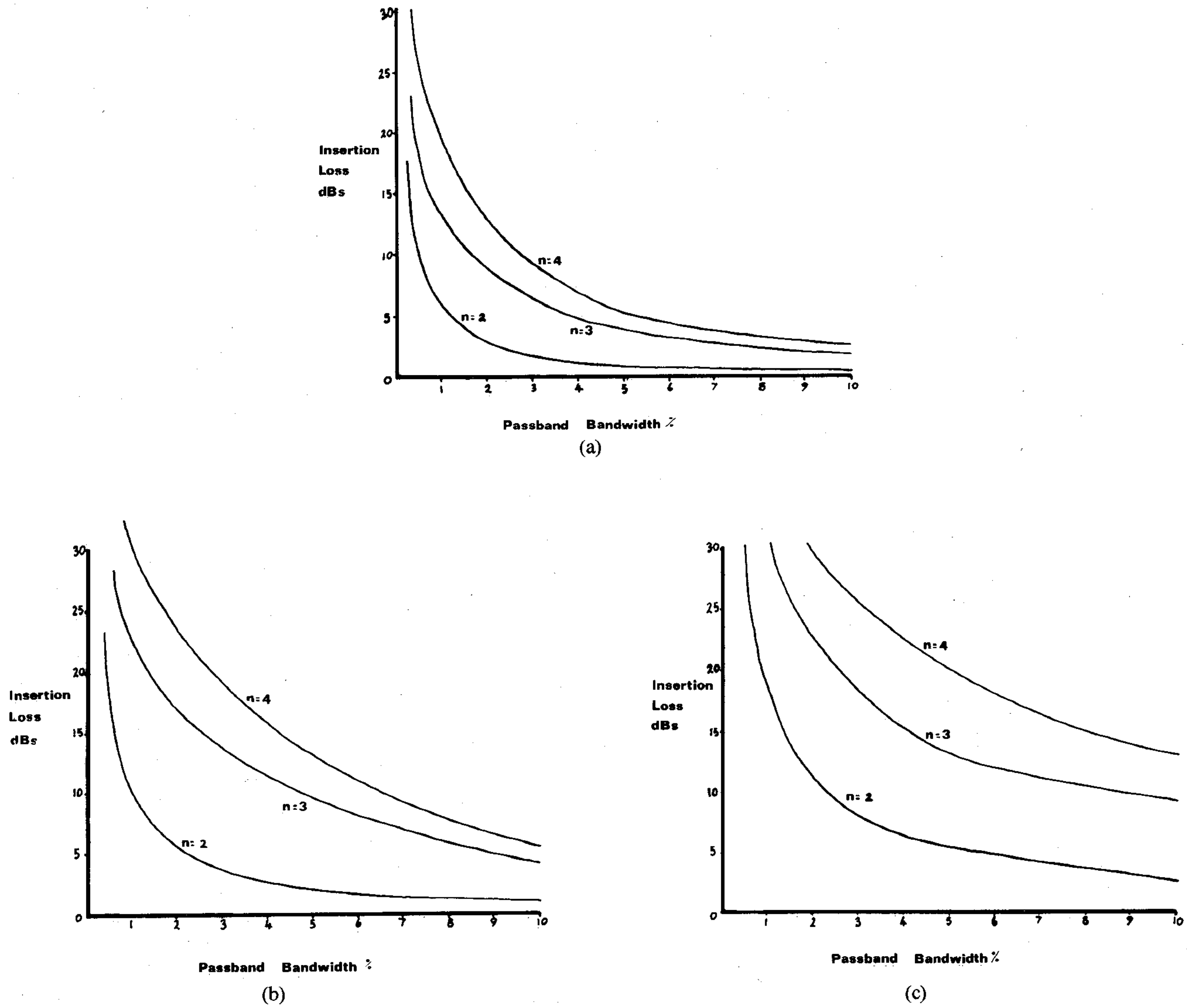

Fig. 8. (a) Zero-bias insertion loss of varactor tuned combline filters as a function of degree passband bandwidth and center frequency. Center frequency 2 GHz. (b) Zero-bias insertion loss of varactor tuned combline filters as a function of degree passband bandwidth and center frequency. Center frequency 4 GHz. (c) Zero-bias insertion loss of varactor tuned combline filters as a function of degree passband bandwidth and center frequency. Center frequency 8 GHz.

in (30) $Y_1 + Y_{12} = 0.8143$. From (31) $\alpha = 19.09$. From (32) $n1 = n2 = 4.666$. From (33) $Y_{12} = 0.0549$. Thus $Y_1 = 0.8143 - 0.0549 = 0.7594$. From (27), (28), and (29)

$$Y_0 = 0.6969$$

$$Y_{01} = 0.3031$$

$$Y_1 = 0.2572.$$

Y_1 must be increased to remove the negative Y_1. Thus $Y_1 = 0.7594 + 0.2572 = 1.0166$ and similarly $Y_2 = 1.0166$.

The filter was constructed in suspended substrate stripline, conversion from electrical to physical parameters was performed using Getsingers method [7]. Grounding of the ends of the combline resonators was achieved by clamping the edge of the stripline circuit board between the upper and lower halves of the stripline housing. Both the circuit

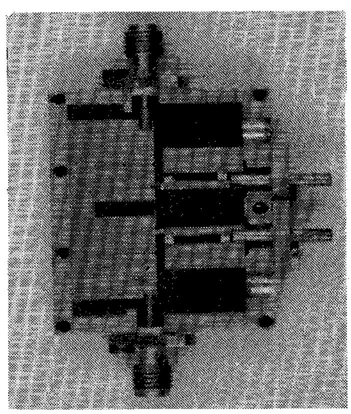

Fig. 9. Picture of the experimental varactor tuned filter.

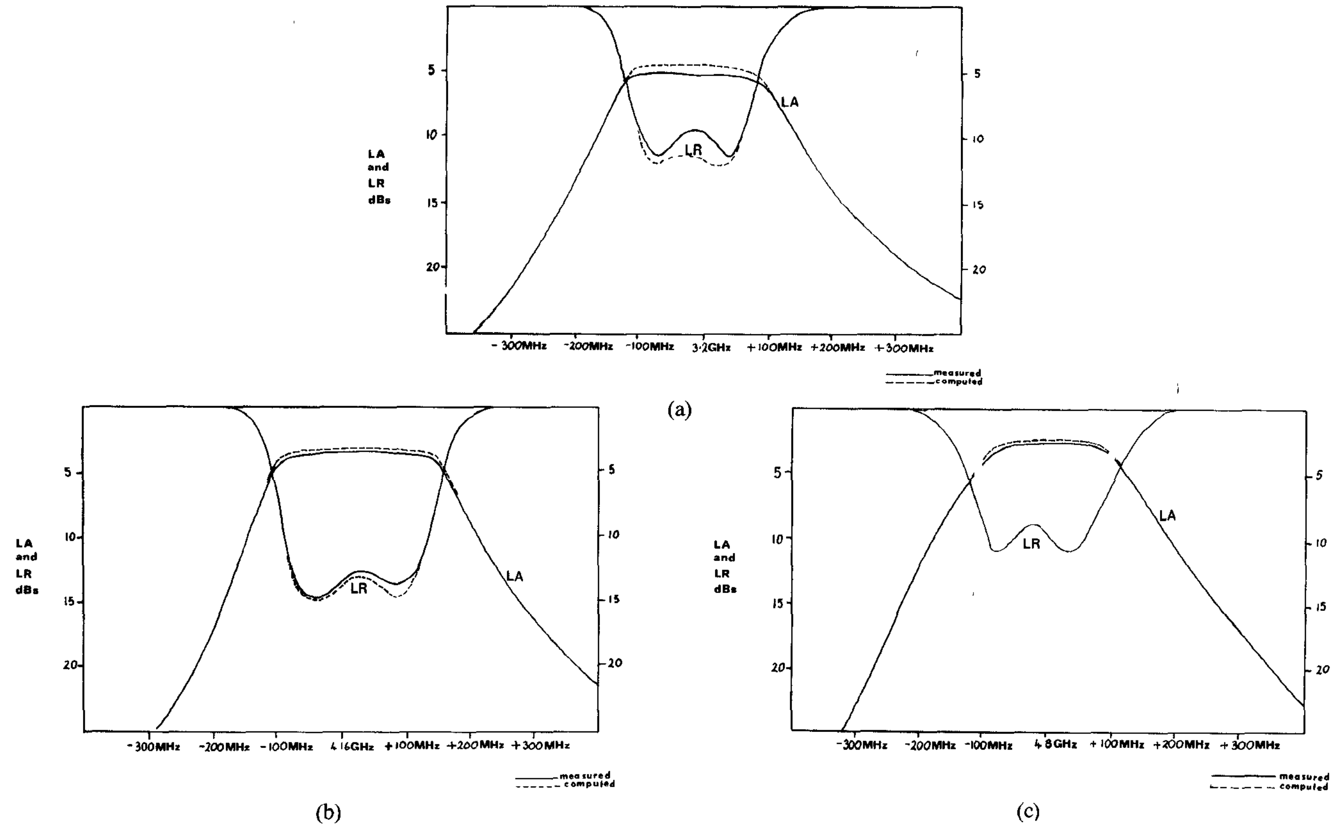

Fig. 10. (a) Frequency response of varactor tuned combline filter at zero bias. (b) Frequency response of varactor tuned combline filter with 4-V bias. (c) Frequency response of varactor tuned combline filter with 30-V bias.

board and the housing were gold plated to assist in achieving a good ground connection. Bias filters for the varactors were realized using printed quarter-wave transmission lines shunted at both ends by 10-pF chip capacitors. A picture of the interior of the filter is shown in Fig. 9.

The experimental performance of the varactor tuned filter is shown for three values of tuned frequency in Fig. 10(a)–(c). In summary, the filter maintained 8-dB passband return loss across a 1700-MHz tuning band. The maximum passband insertion loss was 5.4 dB and the passband bandwidth variation across the tuning band was 12.3 percent. The center frequency of the tuning band was 340 MHz lower than the design aim. This was caused by the varactor bondwire inductance and can be compensated for by shortening the lengths of the filter resonators.

Large signal measurements of the varactor tuned filter indicated a second ordered intercept point of 7 dBm at zero bias and 31 dBm at 30-V bias.

IV. Conclusions

A novel design technique for varactor tuned combline filters has been developed. This design technique enables filters to be tuned across broad bandwidths with almost constant passband bandwidth and return loss. An experimental filter has been constructed in suspended substrate stripline and it exhibits excellent agreement with theory.

Appendix
Derivation of the Element Values of the Transformer Elements in the Tunable Combline Filter

The input/output coupling network is shown in Fig. 5. It will now be assumed, for simplicity, that the network is symmetrical, i.e., $Y_0 = Y_1$. Evaluating the input admittance of this network, looking back into the load via Y_1 and after scaling by the factor $\tan(a\omega)/\tan(a\omega_0)$, we obtain the following real and imaginary parts:

$$\mathbb{R}\mathrm{e}\, Y_{in}(j\omega) = \frac{\tan(a\omega)}{\tan(a\omega_0)\left((1+Y_0/Y_{01})^2+\tan^2(a\omega)/Y_{01}^2\right)} \tag{A.1}$$

$$\mathbb{I}\mathrm{m}\, Y_{in}(j\omega) = \frac{-\left(Y_0(1+Y_0/Y_{01})(2+Y_0/Y_{01})+\tan^2(a\omega)(1+Y_0/Y_{01})/Y_{01}\right)}{\tan(a\omega_0)(1+Y_0/Y_{01})^2+\tan^2(a\omega)/Y_{01}^2}. \tag{A.2}$$

For a perfect match at $\omega = \omega_0$ we require

$$\mathbb{R}\mathrm{e}\, Y_{in}(j\omega_0) \equiv 1 \tag{A.3}$$

$$\mathbb{I}\mathrm{m}\, Y_{in}(j\omega_0) \equiv 0. \tag{A.4}$$

From (A1) and (A3) we obtain

$$(1 + Y_0/Y_{01})^2 + \tan^2(a\omega_0)/Y_{01}^2 = 1 \tag{A.5}$$

or

$$Y_0^2 + 2Y_0Y_{01} + \tan^2(a\omega_0) = 0. \tag{A.6}$$

Substituting (A.6) into (A.1) we obtain

$$\mathbb{R}\mathrm{e}\, Y_{in}(j\omega) = \frac{\tan(a\omega)}{\tan(a\omega_0)\left(1 - \frac{\tan^2(a\omega)}{Y_{01}^2} + \frac{\tan^2(a\omega_0)}{Y_{01}^2}\right)}. \tag{A.7}$$

Now with $a\omega = \theta$, and after some simple manipulation, we obtain

$$\mathbb{R}\mathrm{e}\, Y_{in}(j\omega) = \frac{Y_{01}^2 \sin(\theta)\cos(\theta)\cos^3(\theta_0)}{\sin(\theta_0)\left(Y_{01}^2\cos^2(\theta)\cos^2(\theta_0) + \cos^2(\theta_0) - \cos^2(\theta)\right)}. \tag{A.8}$$

Examining (A.8) we see that if the denominator were forced to be frequency invarient, then we would obtain the required slow frequency variation of $\sin(\theta)\cos(\theta) = \sin(2\theta)/2$. Thus, to remove the denominator frequency dependence we have

$$\underline{\underline{Y_{01} = 1/\cos(\theta_0)}} \tag{A.9}$$

and

$$\underline{\underline{\mathbb{R}\mathrm{e}\, Y_{in}(j\omega) = \frac{\sin(2\theta)}{\sin(2\theta_0)}}}. \tag{A.10}$$

Substituting (A.9) into (A.6) we obtain

$$\underline{\underline{Y_0 = 1 - 1/\cos(\theta_0)}}. \tag{A.11}$$

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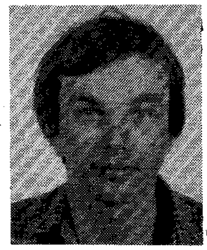

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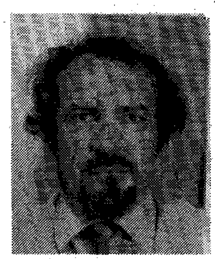

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