Stepped-Ferrite Tunable Evanescent Filters

RICHARD V. SNYDER, MEMBER, IEEE

Abstract—A new technique is described for the design of magnetically tunable filters. With this approach, the resonant sections tune at the same rate enabling maintenance of response shape as center frequency is varied. The filters are minimum-phase structures, realizable with as many poles as required. The design is capable of handling much greater power than single-crystal YIG designs, although the tuning bandwidth is much less.

I. Introduction

THIS PAPER will describe a ferrite tunable bandpass filter constructed in resonated below-cutoff waveguide. The particular example described herein is a 5-pole design tuning the frequency range 8.4–9.2 GHz. A novel method is used to enable the resonated sections to tune at the same rate, thus achieving good filter performance over a wide frequency range. Filters utilizing this principle of operation can be built with arbitrary selectivity (i.e., with 5, 6, 7, or more poles) and may thus complement the YIG filter technology, when the requirements stipulate more than 4-pole selectivity combined with restricted tuning range and moderate insertion loss (i.e., radar, troposcatter, or other communication applications rather than ECM). It is expected that the structure will exhibit a higher power capability than a YIG filter due to the greater limiting threshold.

II. Description

The basic filter is illustrated in Fig. 1. The basic filter synthesis equations are presented in the Appendix (from [3]). Fixed capacitive screws are used to resonate sections of below-cutoff waveguide. The waveguide sections are symmetrically loaded with ferrite slabs. Magnetic tuning is used to vary the cutoff frequencies of each section, thus tuning the center frequency of the composite structure. A different cutoff frequency is used for each length of waveguide interposed between adjacent susceptive parts. The differing cutoff frequencies are achieved by stepping the ferrite slabs. Thus, it is possible to tune each section at the resonant frequency versus applied magnetic field slope sufficient to correct for the loading of the adjacent sections. The net result is the achievement of a synchronously tuned filter, with an acceptably constant bandwidth, over a 10-percent center frequency tuning range.

The structure is symmetrical and is magnetized symmetrically, resulting in the reciprocity necessary for operation as a filter. Fig. 2 depicts the equivalent circuit and summarizes the relationships which describe the structure.

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The author was with Premier Microwave Corp., Port Chester, NY 10573. He is now with R S Microwave Co., Butler, NJ, 07405.

In this paper, no attempt is made to include the discontinuity effects at the junction of two adjacent ferrite loaded sections, differing in cutoff frequency.

Heuristically, for filters of instantaneous bandwidth 1–5 percent, the ferrite sections have lengths at least 5 times the height of the waveguide. Experimentally, discontinuity effects appear to be relatively unimportant. A more accurate network model would treat the junctions in the manner of [6], [7], and [8] in which "accessible" and "localized" modes are considered. However, this model would be most useful for synthesis of large instantaneous bandwidth filters in which the interconnecting ferrite sections are short. This is not a likely application for the new structure due to the somewhat restricted tuning range.

III. Cutoff Frequency for Two-Slab Loaded Rectangular Waveguide

An expression for the propagation constant for the 2 ferrite slab loaded infinite rectangular waveguide is derived in [2]. It is assumed that the slabs are long enough so that the ends do not interact. The nomenclature is as that utilized in [1] and is summarized below. Under those circumstances the propagation constant of the structure is given by

$$\tan k_a(L-2\delta) = \frac{\frac{2k_f\mu_0}{\mu_e k_a}\cosh k_a\delta \sinh k_f\delta}{\sinh^2 k_f\left\{1-\left(\frac{\beta\mu_0 k}{j\mu_e\mu k_a}\right)^2\right\}-\left(\frac{\mu_0 k_f}{\mu_e k_a}\right)\cosh^2 k_f\delta} \tag{1}$$

where

L unloaded waveguide width,

δ ferrite slab thickness,

$k_a^2 = \omega^2\mu_0\epsilon_0 - \beta^2$

$k_f^2 = -\omega^2\mu_e\epsilon_0\epsilon_r + \beta^2$

$\mu_e = (\mu^2 - k^2)/\mu$

ϵ_r ferrite relative dielectric constant,

$k = \rho/(1-\sigma^2)$

$\mu = 1-\rho\sigma/(1-\sigma^2)$

$\rho = 2.8/\omega(4\pi M_s)$

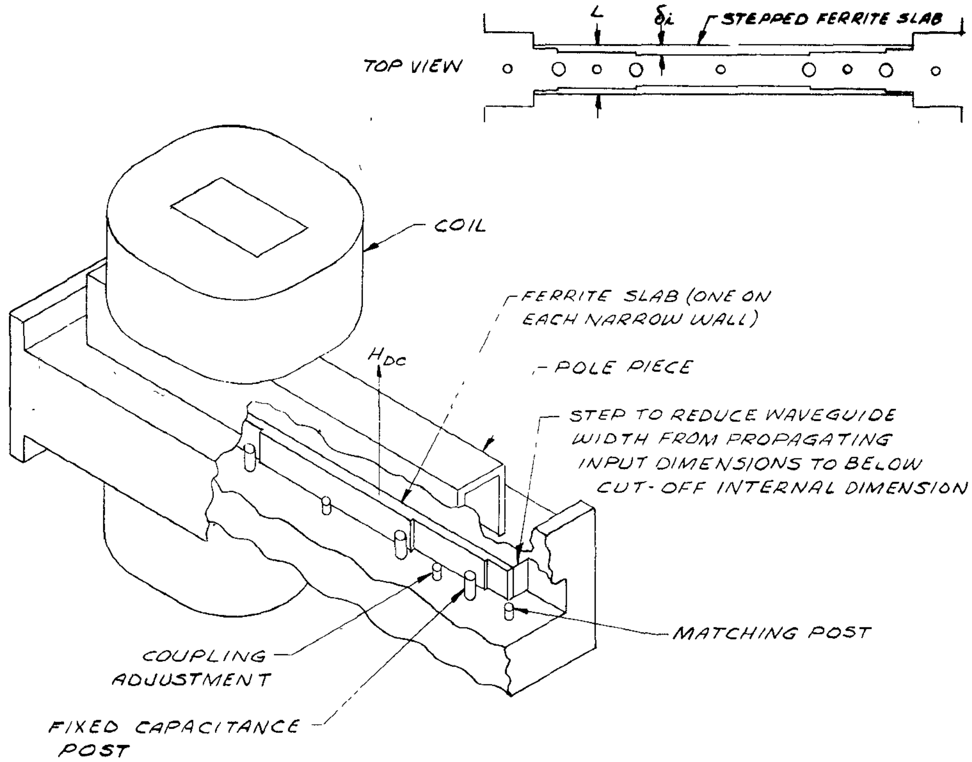
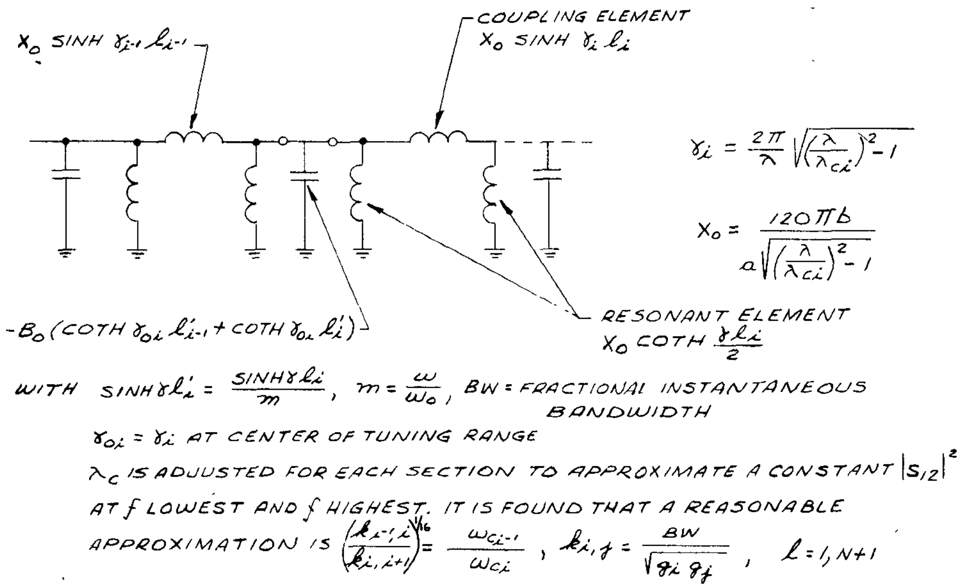
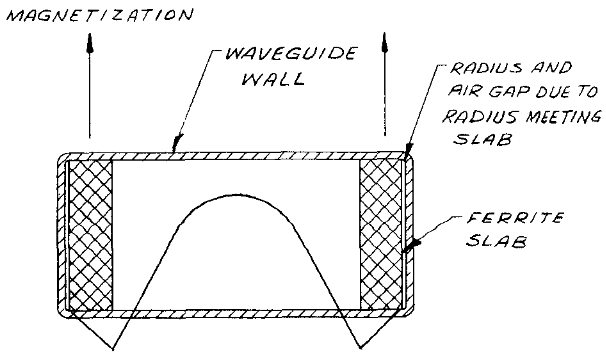


Fig. 1. Basic ferrite tunable filter.



(a)



(b)

Fig. 2. (a) Tunable filter network using fixed capacitors. (b) Electric field distribution of ferrite loaded waveguide mode used for filters.

$\sigma \quad = 2.8/\omega H_t$

H_t dc magnetic field within the ferrite, in oersteds.

Under magnetized conditions, operation is analogous to operating within an unloaded waveguide with an "a" dimension nearly that of the unloaded below cutoff guide minus twice the slab thickness. Thus, propagation is evanescent within the slabs, i.e., $B^2 < 0$.

For a real value for K_f, $\beta^2 < 0$ requires a negative value for μ_e. Note that the $\mu = 0$ condition (i.e., resonance) is to be avoided. For K_f imaginary, $\beta^2 < 0$ requires a positive value for μ_e. The cutoff frequency is obtained, for the fundamental mode, by setting $\beta = 0$. The resulting equation is similar to (4) of [2]

$$\frac{\mu_e}{\mu_0\epsilon_r} + \tanh^2\left(\omega_c(\mu_e\epsilon_0\epsilon_r)^{1/2}\delta\right) = \frac{1+\cos\left\{\omega_c(\mu_0\epsilon_0)^{1/2}(L-2\delta)\right\}}{1-\cos\left\{\omega_c(\mu_0\epsilon_0)^{1/2}(L-2\delta)\right\}}. \tag{2}$$

The applicable root of (2) can be found more or less successfully for a given value of μ_e using various search methods, if one starts close to the value established by the approximations in Section IV. However, the range of variation is great, causing some difficulty in locating roots if the tolerance and starting search frequency are not carefully chosen. The cutoff frequencies are decreased when μ_e is increasingly positive and increased when μ_e is increasingly negative.

The cutoff frequencies for various slab thicknesses can be calculated from (2) of this paper. It was found that agreement between calculated and measured cutoff was strongly dependent upon the corner radius of the waveguide. The results for two values of corner radius are contained in Table I. The reason for the large sensitivity to ferrite fill is not totally clear, but consideration of Fig. 2(b) (from [4]) indicates that cutoff should be at least dependent upon the presence or absence of air gaps near the ferrite slabs. Therefore, it has been found desireable to initially measure the available cutoff variation of a short length of ferrite-loaded guide under the same magnetization conditions intended for use on the filter. It is probable that the actual values for μ_e range from small positive to small negative, in practice. To give some feeling for the calculated cutoff frequency f_c and resonant frequency f_r consider the following Table II.

TABLE I

δ	f_c (GHz) $R < .005$	f_c (GHz) $R \cong .03$
.08	11.2 - 12.7	9.75-11.3
.083	11.24- 12.77	9.78-11.4
.085	11.31-12.9	9.83-11.49

MEASURED CUT-OFF FREQUENCY f_c
TUNABLE RANGE (GHz), $0 < H_{dc} < 1500$
FOR 2 VALUES OF WAVEGUIDE
CORNER RADIUS R

TABLE II
FILTER SECTION MEASURED RESONANT FREQUENCY AND CALCULATED FERRITE CUTOFF FREQUENCY

μ_e	δ	Resultant H_i (calc. from Fresonant)	Filter Section Resonant Freq. f_r (GHz)	Calculated f_c (GHz)
.300	..08	400	9.0	11.819
.300	.083	400	9.055	11.949
.300	.085	400	9.12	12.039
.100	.08	780	9.0	12.378
.100	.083	780	9.07	12.524
.100	.085	780	9.15	12.629
-.100	.080	1010	9.0	12.763
-.100	.083	1010	9.09	12.895
-.100	.085	1010	9.19	13.048
-.200	.080	1125	9.0	12.790
-.200	.083	1125	9.09	12.934
-.200	.085	1125	9.19	13.049

With H_i from basic lossless ferrite tensor component equations, Sec. III and with f_c calculated using (2)

$\epsilon_r = 12$

$4\pi M_s = 2300$ (MMF-10 MATERIAL)

$\delta = .08$ (INSIDE WAVEGUIDE IS .622" x .4")

FEED WAVEGUIDE = WR-90 (.9" x .4")

IV. Design Technique and Example

To tune in frequency from an initial frequency f_0 to a new resonant frequency f_r, with reference to Fig. 1, we require all filter sections to tune at the same rate, thus

$$-B_0(\coth\gamma_0 l_{i'-1} + \coth\gamma_0 l'_i) = B_r(\coth\gamma_r l'_{i-1} + \coth\gamma_r l'_i) \tag{3}$$

i.e., reactance of fixed capacitor, calculated = reactance of variable inductance from synthesis (Appendix).

For constant bandwidth, we require

$$\frac{1}{B_0}\sinh\gamma_0 l_i = \left(\frac{1}{B_r}\sinh\gamma_r l_i\right)\frac{\omega_0}{\omega_r}$$

$$\text{substituting } \gamma_i = \frac{2\pi}{\lambda_i}\sqrt{\left(\frac{\lambda_i}{\lambda_{ci}}\right)^2 - 1}$$

$$B_i = \frac{a}{120\pi b}\sqrt{\left(\frac{\lambda_i}{\lambda_{ci}}\right)^2 - 1} \tag{4}$$

into (3), with the physical lengths l_i fixed and derived from synthesis of the filter at the geometric center of the tuning range. Let us first consider one isolation section.

For any tuned frequency to have the left-hand side of

(3) = right-hand side of (3) we have

$$\sqrt{\left(\frac{\lambda_{0i}}{\lambda_{ci}}\right)^2-1}\coth\left[\frac{2\pi l_i}{\lambda_{0i}}\sqrt{\left(\frac{\lambda_{0i}}{\lambda_{ci}}\right)^2-1}\right]$$

$$=\sqrt{\left(\frac{\lambda_{ri}}{\lambda_{rci}}\right)^2-1}\coth\left[\frac{2\pi l_i}{\lambda_{ri}}\sqrt{\left(\frac{\lambda_{ri}}{\lambda_{rci}}\right)^2-1}\right]$$

with $\lambda_{0i}=T\lambda_{ri}$

$$\sqrt{\left(\frac{T\lambda_{ri}}{\lambda_{ci}}\right)^2-1}\coth\left[\frac{2\pi l_i}{T\lambda_{ri}}\sqrt{\left(\frac{T\lambda_{ri}}{\lambda_{ci}}\right)^2-1}\right]$$

$$=\sqrt{\left(\frac{\lambda_{ri}}{\lambda_{rci}}\right)^2-1}\coth\left[\frac{2\pi l_i}{\lambda_{ri}}\sqrt{\left(\frac{\lambda_{ri}}{\lambda_{rci}}\right)^2-1}\right]. \quad (5)$$

Equation (5) is approximated for small values of T

$$\frac{\lambda_{0c}}{\lambda_{rc}}\cong 1.09\frac{\lambda_0}{\lambda_r}T, \qquad 0.9\leq T\leq 1.1 \quad (6)$$

i.e., cutoff frequency for each single, isolated (i.e., uncoupled) filter section must move in frequency at about 1.090 times the rate of the desired filter center frequency, if synchronous tuning with fixed tuning capacitors is desired. To consider the loading effect of each adjacent section, rewrite (3)

$$-B_0(\coth\gamma_0 l'_{i-1}+\coth\gamma_0 l'_i)$$
$$-B_r(\coth\gamma_r l'_{i-1}+\coth\gamma_r l'_i)=0. \quad (7)$$

Selecting λ_{ci}, λ_{0i}, and λ_{ri} enables finding the roots of (7) in terms of the unknown λ_{cri}, the cutoff frequency of each filter section when the entire filter is tuned *from* $\lambda_{0i}=\lambda_r$ to wavelength $\lambda_{0i}=\lambda_0$. A useful approximation to the root positions is again found to be

$$\frac{\lambda_{0c}}{\lambda_{rc}}\cong 2.2\frac{\lambda_0}{\lambda_r}T, \qquad 0.9\leq T\leq 1.1 \quad (8)$$

i.e., the average cutoff frequency for the composite of *doubly loaded* sections of waveguide comprising the filter must move 2.2 times as far as the corresponding tuned frequency for the filter: substituting each pair of values l_{i-1} and l'_i (for all i lengths) into (7) enables determining the required cutoff frequency variation for each section—considering only the requirement for synchronous tuning.

However, the solutions to (8) will not satisfy (4); thus, requiring perfectly synchronous tuning conflicts with the requirement for the constant bandwidth. As in any tunable filter, compromise is required to provide a reasonable response shape with small bandwidth variation, as the structure is tuned.

Stepping the ferrite, as discussed in Section II provides the section-by-section cutoff frequency variation required. Alternatively, the magnetic field could be adjusted for each section. Even with a shaped pole piece, this has not proven feasible. A word about the approximations (6) and (8). Values have proven most accurate for waveguides near a 2:1 aspect ratio.

For the specified tuning range (8) will provide the average cutoff frequency variation required for the ferrite loaded structure. Values of (μ_e) at the geometric center of the tuning range and selected, and substituted into (2) to find the mean loaded waveguide cutoff frequency. The values of μ_e are selected by considering the available ferrite material and magnetic bias, using (1).

Using [3] (see Appendix), the bandpass filter is synthesized at the geometric center of the specified tuning range. A reasonable approximation to satisfaction of (4) and (5) (i.e., synchronous tuning and constant bandwidth) is obtained by requiring the calculated cutoff frequency for each section (calculated at the geometric center of the required tunable range) to vary as follows:

$$\frac{f_{cij}}{f_{cjk}}=\left(\frac{K_{ij}}{K_{jk}}\right)^{1/16},$$

$$K_{ij}=\frac{W}{\sqrt{g_i g_j}}, \qquad W=\text{fractional bandwidth.}$$

The interstage couplings determine the total excursion for the connecting sections which comprise the series coupling inductive elements (see Fig. 1). One must ensure that bias is sufficiently below the resonant ($\mu=0$) condition. Failure to consider the magnetic operating range will result in excess insertion loss. Sensitivity of the cutoff frequency to ferrite thickness is such that only small steps (0.002 − 0.005) are found to be required on the ferrite slabs for the X-band example described herein.

Temperature compensation of the structure can be accomplished either on a magnet compensation or dc current variation basis. Considering that the $4\pi M_s$ of the ferrite material decreases with increasing temperature and that the filter resonant frequency will thus also decrease, it is necessary to supply additional magnetic bias to compensate. The magnet may be temperature compensated by placing in shunt a special steel whose permeability decreases with increasing temperature. Thus, as temperature increases, a magnetic field redistribution will occur in the desired direction. Alternatively, or simultaneously, a thermistor circuit can be used to adjust the dc current flow to the magnet, under constant voltage excitation.

A sample 5-pole design was constructed using the measured cutoff data of Table I ($R=0.03$). A computer program based upon the Appendix and approximation (8) was written and used for the synthesis. (Fig. 3 is a sample printout). The program enables determination of the cutoff variation which preserves the physically fixed lengths of waveguide. Performance data is summarized in Figs. 4–7. The instantaneous bandwidth of the device shows greater variation than was originally expected. Adjustment of slab step thickness resulted in some improvement. Adjustments were of the order of 0.0002 in. A computer program incor-

```
0
INP. 0 FOR RECT,1 FOR ROUND
?
 0
EVANESCENT TE10 MINIMUM PHASE B.P. FILTER
ENTER 0 FOR SHUNT CASE,1 FOR SERIES
?
 1
ENTER N,RIP,0(GCOMP)1(GREAD),Z0,A,B,D,DD,F0,F1,F2
?
 5,.04,0,.50,.,9,.4,.605,.4,8.4E09,8.31E09,8.49E09
ELEMENT VALUES ARE

 1 =  0.957600

 2 =  1.370386

 3 =  1.787542

 4 =  1.370388

 5 =  0.957607

 6 =  1.000000
THEORETICAL QU FOR COPPER IS    0.6868426E+04
TUNING CAPACITANCES ARE
 0.4506230E-13
 0.9003811E-13
 0.8995170E-13
 0.9003811E-13
 0.4506230E-13
THE LENGTHS ARE=(INCHES)
0.271658
1.170587
1.288626
1.288626
1.170589
0.271658
TOTAL LENGTH IS  5.461742
OUTPUT SHUNT REACTANCE,PROP.GUIDE Z0,TERM. RES.        0.0          155.075897
  426.664062       0.086950
BP SPURIOUS FREQS ARE(GHZ)
     23.813507
     10.980335
     10.776112
     10.776112
CAP.DESIGN?TYPE 1.,NO=0.
```

$f_0 = 8.4$
$BW = 190$
$rip = .04$
$f_c = 9.754$

EQUAL CAPACITORS TO UNIT TUNED AT 9.2 GHz

END SECTIONS TRACK MIDDLE SECTIONS REQUIRE DIFFERENT RATE OF CUT-OFF VARIATIONS

```
 1
ENTER N,RIP,0(GCOMP)1(GREAD),Z0,A,B,D,DD,F0,F1,F2
?
 5,.1,0,.50,.,9,.4,.51,.4,9.2E09,9.1E09,9.3E09
ELEMENT VALUES ARE

 1 =  1.146837

 2 =  1.371209

 3 =  1.975029

 4 =  1.371210

 5 =  1.146844

 6 =  1.000000
THEORETICAL QU FOR COPPER IS    0.6866641E+04
TUNING CAPACITANCES ARE
 0.4472786E-13
 0.8941676E-13
 0.8937732E-13
 0.8941676E-13
 0.4472786E-13
THE LENGTHS ARE=(INCHES)
0.273030
0.919476
0.992161
0.992161
0.919478
0.273030
TOTAL LENGTH IS  4.369334
OUTPUT SHUNT REACTANCE,PROP.GUIDE Z0,TERM. RES.        0.0          132.487885
  380.177246       0.117900
BP SPURIOUS FREQS ARE(GHZ)
     24.517410
     13.232328
     13.010822
     13.010817
CAP.DESIGN?TYPE 1.,NO=0.
?
 0
INP. 0 FOR RECT,1 FOR ROUND
?
 0
EVANESCENT TE10 MINIMUM PHASE B.P. FILTER
ENTER 0 FOR SHUNT CASE,1 FOR SERIES
```

$f_0 = 9.2$
$BW = 200$
$rip = .1$
$f_c = 11.571$ GHz

EQUAL CAPACITORS TO UNIT TUNED AT 8.4 GHz

Fig. 3. Varying cutoff to maintain constant length in constant capacitance tunable filters.

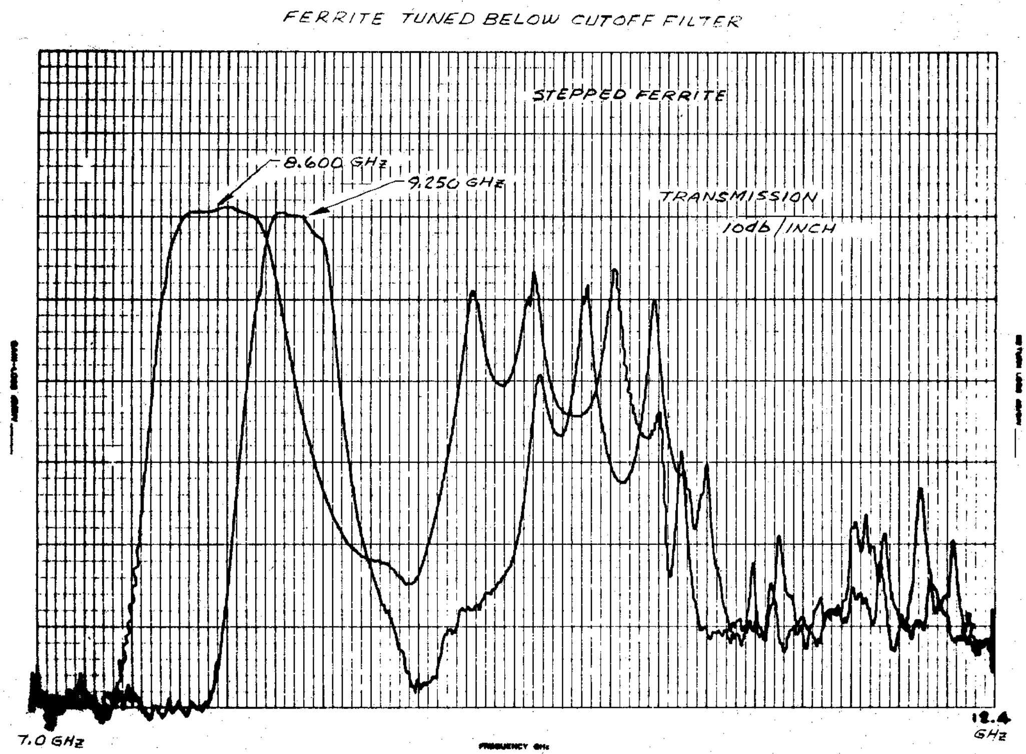


Fig. 4. Response characteristics.

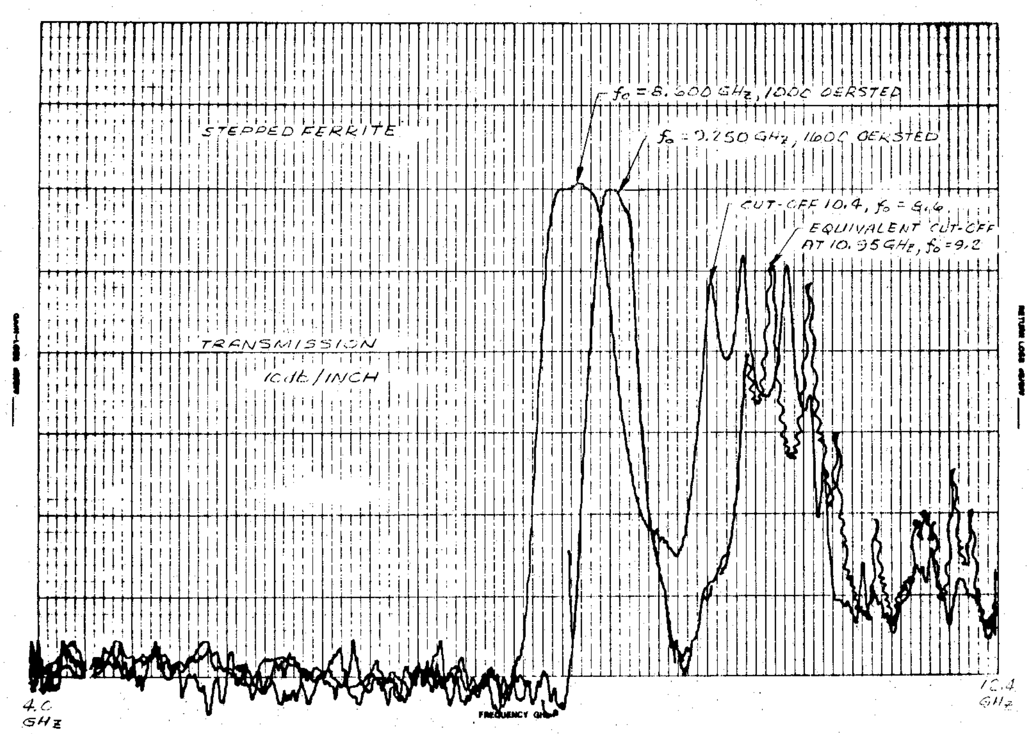


Fig. 5. Response characteristics.

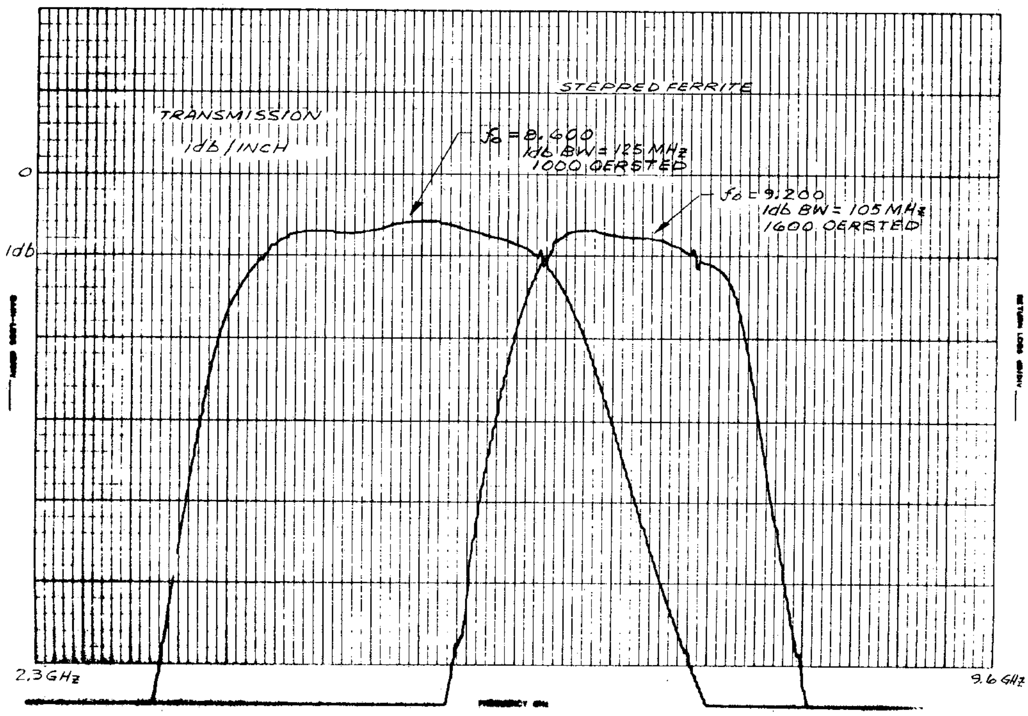


Fig. 6. Response characteristics.

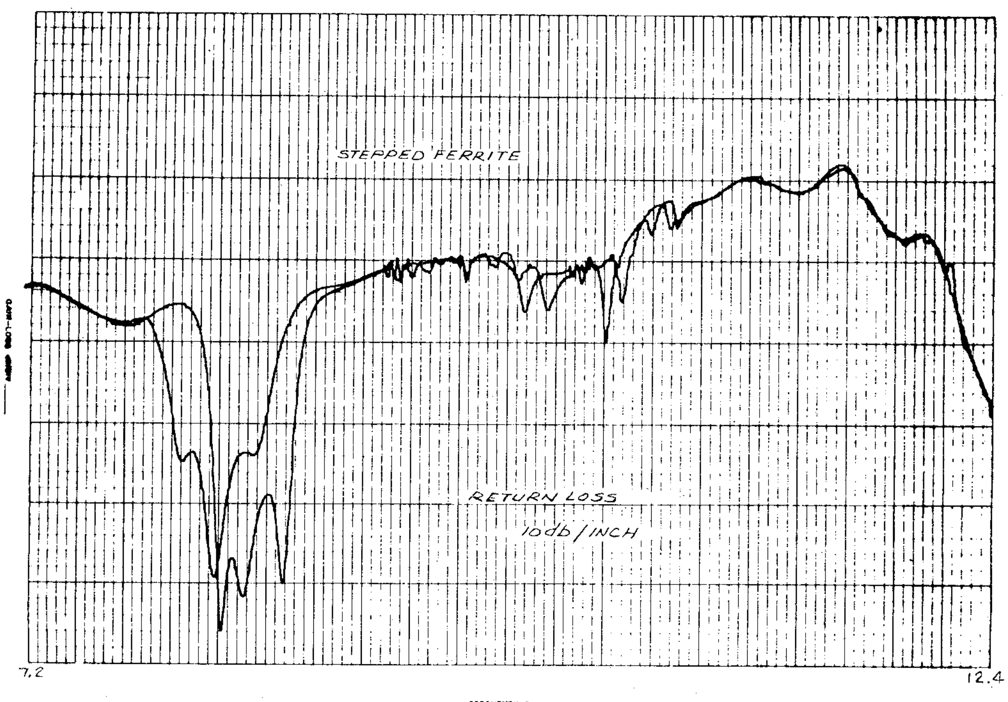


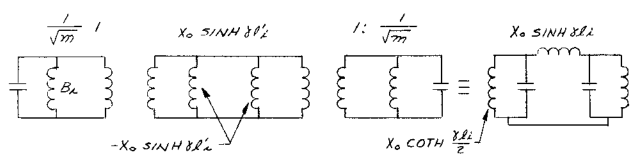
Fig. 7. Response characteristics.

porating (4) and (5) would result in more precise prediction of the section-by-section cutoff variation required. Table III contains temperature data taken on the sample design with and without magnet compensation.

VI. Conclusion

Using stepped ferrite-loaded below-cutoff sections of waveguide, magnetically tunable bandpass filters have been built which can tune at least 800 MHz in X-band. The Q_u (inferred from performance data) is about 25 percent of the theoretical equivalent evanescent (no ferrite) bandpass filter. The structure is simple enough to allow filters with arbitrarily high selectivity to be built. Additional work should be done to resolve the difference between calculated and measured cutoff frequencies.

Appendix
Evanescent Filter Design



(1a) $B_i = B_0 (\coth \gamma \ell'_{i-1} + \coth \gamma \ell'_i)$

(1b) $B_i = \frac{B_0}{m} (\coth \gamma \ell_{i-1} + \coth \gamma \ell_i)$

WHERE

(2a) $\sinh \gamma \ell'_i = \dfrac{1}{\dfrac{\sqrt{V_{i,i-1}} - \sqrt{V_{i,i+1}}}{\left(\frac{W'_g}{W}\right)(\Delta)}}$ (SEE EQ 6 & NOTES)

AND

(2b) $\sinh \gamma \ell_i = \dfrac{\sinh \gamma \ell_i}{M}$, V_i CHOSEN AS IN [5] OR APPROXIMATED BY (6a), (6b)

ℓ'_i = CENTER TO CENTER SPACING BETWEEN i^{th} AND $(i+1)^{th}$ CAPACITOR

B_i = SUSCEPTANCE OF i^{th} SHUNT INDUCTANCE (OR i^{th} SHUNT CAPACITANCE USED FOR RESONANCE

(2c) $M = \frac{\omega}{\omega_0}$

(2d) $\Delta = \dfrac{2}{1 + \frac{1}{1-(\lambda_c/\lambda_0)^2}}$

(2e) $\gamma = \frac{2\pi}{\lambda} \sqrt{\left(\frac{\lambda}{\lambda_c}\right)^2 - 1}$

X = CROSS SECTION DEPENDENT (SEE TEXT)

(2f) $W = \frac{\omega_2 - \omega_1}{\omega_0}$

(2g) $W'_g = 4 \phi'_0$

(2h) TABULATED BW $W_g = \left(\frac{W''_g}{W'_g}\right)(2W)$

WHERE $W''_g = \frac{4\theta'_0}{\pi}$

$\theta'_0 = \sin^{-1}\left(\frac{\omega_0}{\omega_1} \sin \pi \frac{\omega_1}{\omega_0}\right) = \sin^{-1}\left(-\frac{\omega_0}{\omega_2} \sin \pi \frac{\omega_2}{\omega_0}\right)$

(3) $\sinh \phi'_0 = \frac{\omega_0}{\omega_1} \sinh\left(\omega_0 - \frac{\omega_1}{\omega_0}\right) = -\frac{\omega_2}{\omega_0} \sinh\left(\omega_0 - \frac{\omega_2}{\omega_0}\right)$

INSERTION LOSS FUNCTION

(4) $\frac{P_0}{P_L} = 1 + h^2 Th^2 \left[\frac{\omega_0}{\omega} \frac{\sinh\left(\omega_0 - \frac{\omega}{\omega_0}\right)}{\sinh \phi'_0} \right]$

RIPPLE $= 10 \log (1 + h^2)$

(5) $h = \frac{V-1}{2\sqrt{V}}$ V = MAXIMUM VSWR FOR NARROW BAND ONLY

(6a) $V_{0,1} = \frac{2}{\pi} \frac{g_1}{W} = V_{N,N+1}$ WHERE $W = \frac{W'_g}{2}$

(6b) $V_{i,i+1} = \frac{4}{\pi^2} \frac{g_i g_{i+1}}{W^2}$ $i = 1, 2, \ (N-1)$

TABLE III
Temperature Data

	T(°C)	I_{DC}	f_o (MHz)
NO COMPENSATION	25°	1 A	8415
	85°	1 A	8375
	85°	1.1A	8415
PARALLEL SHUNT ON POLE PIECES	25°	1 A	8415
USING CARPENTER 30	85°	1 A	8405
TEMP. COMPENSATING ALLOY	85°	1.02A	8415

Table III contains temperature data taken on the sample design with and without magnet compensation.

OTHERWISE, USE HALF WAVE FILTER OR QUARTER-WAVE TRANSFORMER PROTOTYPE TABLES FOR VALUES OF $V_{i,j}$, AS IN [5]. PUBLISHED PROTOTYPE BANDWIDTHS W_g MAY BE USED IF ADJUSTED

PROCEDURE

1) COMPUTE ϕ'_0 USING (3)
2) SELECT N USING (4) TO PREDICT STOPBAND RESPONSE
(3) COMPUTE W'_g USING (2g)
(4) COMPUTE LENGTHS USING (2a) & (6) OR (2a), (2h) & TABLES
(5) COMPUTE CAPACITORS USING (1a)
(6) COMPUTE RESPONSE AND OPTIMIZE

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