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BROADBAND WAVEGUIDE FILTERS WITH WIDE STOPBANDS USING A STEPPED-WALL EVANESCENT MODE APPROACH

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ABSTRACT

Utilizing more than one below cut-off cross section, evanescent mode series coupled bandpass filters are described which achieve wide passbands and stopbands with practical mechanical dimensions. Performance is comparable to the conventional approach. The new design is applicable for center frequencies to above 40 GHz.

Introduction

Filter applications currently growing in importance require broadband (10% and up) waveguide bandpass filters with broad stopbands. These filters are utilized as preselectors or in multiplexers and are frequently required to cover up to full waveguide bandwidths with stopbands extended to the 3rd harmonic or above.

Approaches to the construction of broadband waveguide filters include the E-plane or fin line filter [4], [5], inductive iris or inductive post design [3], [6], or the conventional evanescent waveguide design [1], [2]. The E-plane approach is quite inexpensive and provides wide stopbands but is very difficult to realize for bandwidths in excess of 10% (see Fig. 1). Inductive iris filters become impractical due to iris size, interaction and stopband flare above 20%, while post filters do not achieve wide stopbands (typically re-entering near the 2nd harmonic). Conventional evanescent designs are capable of alleviating all difficulties enumerated above, but suffer from close spacing of the first few capacitive pins. This shortcoming is intrinsic to the structure, for bandwidths in excess of 30%. Because this mechanical problem adversely affects tuning time (and hence economics), we have developed a modified design approach for evanescent filters, which eliminates the close spacing problem.

Approach

The theory of the series coupled evanescent dominant-mode design for filters is reviewed in the Appendix. Fig. 2 illustrates the cross section of a typical conventional series coupled design. Fig. 3 depicts a modified (stepped-wall) design. The end sections use a cross section with a lower cut-off frequency, thus increasing the rod spacing. The filter still maintains the wide stopbands which result from use of a small, far below cut-off cross section, as the majority of the filter is built in the smallest cross section. Fig. 4 defines the network elements for the series coupled evanescent filter. Realization of the new design involves two main problems:

1) Compensation due to susceptive effects at the junction of two cut-off sections of differing cross sections.

2) Bandwidth shrinkage due to finite inductance of the capacitive pins.

Solution of problem (1) is facilitated by deriving expressions similar to (7) and (8) of [1], with both guides cut-off. The resultant expression, which is a good approximation to the susceptance generated at the junction of two cut-off waveguides, is given by (7) and (8) in the Appendix. Fig. 5 defines the junction dimensions and transformer ratios. Problem (2) is solved as is [1], by forming an inductive tee-equivalent to the pin series inductance, imbedding the series arms of the tee into the filter series inductive coupling elements and increasing the resonating capacitance value to compensate for the tee shunt inductance. The interconnecting lengths between pins straddling cross section interfaces are very close to the average of the spacings (computed as in [1]) for each individual cross section. This is true because the discontinuity susceptance due to the junction is small, for small differences in cross section.

To illustrate the design, Table I presents the spacings for the first 4 sections of a filter covering the 18-26 GHz band. Case 1 depicts a filter with a homogenous internal cross section measuring .15"x.15". Case 2 uses a constant .21"x.21". Case 3 utilizes a .21"x.15" cross section for the first 2 sections only, and .15"x.15" for all center sections. Note that the stepped design uses constant rod diameters (best choice for diameter is 20% of the cross section "a" dimension). However, constant diameter is convenient. Spurious passband characteristics are not significantly degraded as compared to the smallest homogenous unit. Again, this is expected as most of the filter sections are contained in the smallest portion. Computed and measured responses of some actual filters are shown in Fig. 6.

Conclusion

The modified evanescent design results in units that are easy to tune, applicable up to full waveguide bandwidths, and practical well above K_a band. Structurally simple, this approach offers compact construction and high reliability. Finally, the stepped wall can be used to provide an end section transformation for coax connection, facilitating the design of frequency selective waveguide to coax adapters.

References

1) R. V. Snyder, "New Application of Evanescent Mode Waveguide to Filter Design", IEEE Trans. MTT, Vol. MTT-25, December 1977.

2) G. Craven and C. Mok, "The Design of Evanescent Waveguide Bandpass Filter for a Prescribed Insertion Loss Characteristic", IEEE Trans. MTT, Vol. MTT-19, March 1971.

3) R. Levy, "Theory of Direct-Coupled Cavity Filters", IEE Trans. MTT, June 1967.

4) Y. Konishi, "Microwave Circuits Constructed Inside a Waveguide", U.S. Pat. 3914713, October 1975.

5) J. Reindel, "Printed WG Circuits Trim Component Costs", Microwaves, Oct. 1980.

6) J. Bratherton, "Waveguide Filters for MM Wavelengths", Microwave Journal, July 1982.

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

(1b) $B_i = \frac{B_x}{m}(\coth \gamma l_{i-1} + \coth \gamma l_i)$

WHERE

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

AND

(2b) $\sinh \gamma l_i' = \frac{\sinh \gamma l_i}{M}$; V_i CHOSEN AS IN [2] OR APPROXIMATED BY (6a), (6b)

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

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

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

X_o = CROSS-SECTION DEPENDENT (SEE TEXT)

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

(2g) $W_q' = 4\phi_o'$

(2h) TABULATED BW

$W_q = \left(\frac{W_q''}{W_q'}\right)(2W)$

WHERE

$W_q'' = \frac{4\theta_o'}{\pi}$

$\theta_o' = \sin^{-1}\left(\frac{\omega_o}{\omega_1}\sin\pi\frac{\omega_1}{\omega_o}\right) = \sin^{-1}\left(\frac{\omega_o}{\omega_2}\sin\pi\frac{\omega_2}{\omega_o}\right)$

(3) $\sinh \phi_o' = \frac{\omega_o}{\omega_1}\sinh\left(\omega_o - \frac{\omega_1}{\omega_o}\right) = -\frac{\omega_2}{\omega_o}\sinh\left(\omega_o - \frac{\omega_2}{\omega_o}\right)$

(a) INSERTION LOSS FUNCTION

(4) $\frac{P_o}{P_L} = 1 + h^2 T_N^2\left[\frac{\omega_o}{\omega}\frac{\sinh\left(\omega_o - \frac{\omega}{\omega_o}\right)}{\sinh \phi_o'}\right]$

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

(5) $h = \frac{V-1}{2\sqrt{V}}$

V = MAXIMUM VSWR

FOR NARROW BAND ONLY

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

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

OTHERWISE, USE HALF-WAVE FILTER OR QUARTER-WAVE TRANSFORMER PROTOTYPE TABLES FOR VALUES OF V_{ij}, AS IN [2]. PUBLISHED PROTOTYPE BANDWIDTHS W_q MAY BE USED IF ADJUSTED USING (2h)

PROCEDURE

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

(7) $n_{HE} = \sqrt{\frac{ab}{Dd}}\frac{1-\left(\frac{D}{a}\right)^2}{\frac{a}{a_g}\cos\frac{\pi D}{2a}}$

(8) $B_{HE} = -\frac{1}{\omega n}\left\{\frac{a_g}{a}\cot^2\frac{\pi D}{2a} + \frac{a}{D}n_H^2\left[\frac{b}{a_g}\left\{\left(\frac{\pi}{D}\right)^2 - k^2\right\}\ln\left(\csc\frac{\pi D}{2a}\right) - \frac{a_g}{a}\left(\frac{b}{D}\right)^2\left\{\frac{1}{3} + \frac{1}{2}\left(\frac{d}{b}\right)^2 - \frac{8}{\pi^2}\frac{d}{b}\right\}\right]\right\}$

WHERE

$a_g = \frac{\lambda_o}{\sqrt{1-\left(\frac{\lambda_c}{2a}\right)^2}}$

$n_H = \sqrt{\frac{a}{D}}\frac{1-\left(\frac{D}{a}\right)^2}{\frac{a}{a_g}\cos\frac{\pi D}{2a}}$

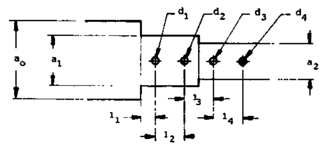
a LARGER GUIDE WIDTH
b LARGER GUIDE HEIGHT
D SMALLER GUIDE WIDTH
d SMALLER GUIDE HEIGHT
λ_c SMALLER GUIDE CUTOFF WAVELENGTH

AND n_{HE} IS THE TRANSFORMATION RATIO OF ADMITTANCE AT THE JUNCTION. THE INPUT SECTIONS OF SUCH FILTERS ARE SHORTENED TO SUBTRACT THE DISCONTINUITY SUSCEPTANCE COMPUTED IN (8) REFERENCE PLANES ARE SHOWN IN FIG 6

APPENDIX

DESIGN EQUATIONS, SERIES COUPLED STEPPED WALL EVANESCENT MODE BANDPASS FILTER

	Case 1 Homogeneous (.15 x .15)	Case 2 Homogeneous (.21 x .21)	Case 3 Stepped Design (.21 x .15, 1st two) (.15 x .15, balance)
a_o	.42	.42	.42
a_1	.15	.21	.21
a_2	.14	.21	.15
d_1	.012	.030	.030
d_2	.020	.030	.030
d_3	.030	.040	.030
d_4	.030	.040	.030
l_1	.008	.019	.019
l_2	.056	.072	.070
l_3	.077	.106	.089
l_4	.082	.115	.113
First spurious freq. for filter (above -50 dbc)	68 GHz	52 GHz	64 GHz

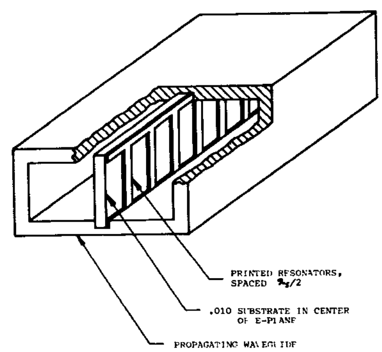


b_o dimension is 0.21 (WR-42 input)

$b_1 = b_2 = 0.15$

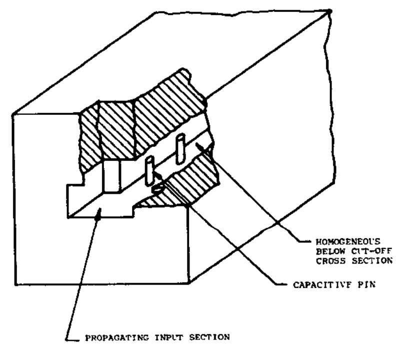
TABLE I

Dimensions for 1st 4 sections of 18-26 GHz filter.



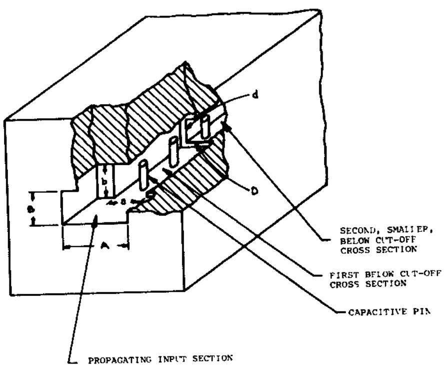
E-PLANE FILTER

FIG. 1



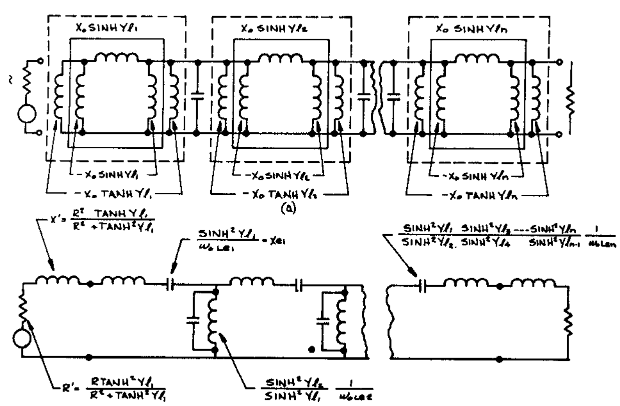
CONVENTIONAL EVANESCENT MODE SERIES COUPLED WAVEGUIDE FILTER

FIG. 2

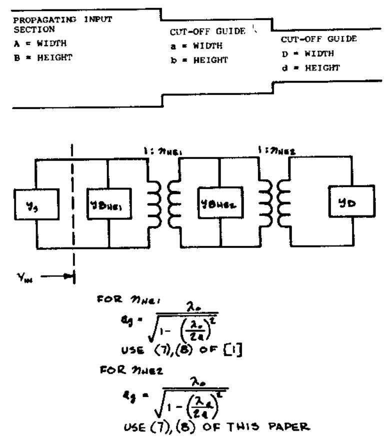


STEPPED WALL EVANESCENT MODE
SERIES COUPLED WAVEGUIDE FILTER

FIG. 3

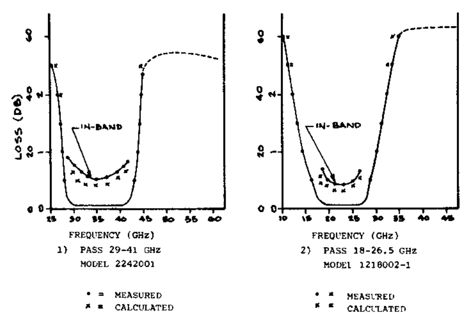


FILTER WITH SERIES RESONATOR TERMINATION
(a) J inverter representation. (b) Derived ladder network.
FIG. 4



WAVEGUIDE STEP DEFINITION OF DIMENSIONS
(SEE EQS. [7], [8])

FIG. 5



RESPONSES OF STEPPED WALL
EVANESCENT BANDPASS FILTERS

FIG. 6