BANDPASS FILTERS USING STRIP LINE TECHNIQUES

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Introduction

Strip lines provide a convenient transmission medium for the realization of microwave filters. Since bandpass filters designed in waveguide and coaxial lines would be large at ultra-high frequencies, strip lines afford a practical means of realizing filters which are simply fabricated, are readily reproduced, and, in most cases, represent an appreciable savings in size and weight. Of the different types of strip transmission lines currently in use, the so-called "sandwich"[2] structure has been employed at Melpar for two reasons: (1) very broad-band coax-to-strip-line transitions are easily realized; and (2) the electromagnetic field is essentially confined between the two ground planes, thus reducing problems in packaging.

Using design techniques [11, 13, 15] developed for direct-coupled cavity-type waveguide and coaxial filters, experimental strip-line filters having ten percent bandwidths in the u-h-f spectrum have been developed. These units have less than 1 db mid-band insertion loss and provide a rejection of greater than 40 db at frequencies twelve percent from the center frequency. The design techniques discussed in this paper are general and, therefore, are not restricted to the realization of the above filter characteristics. Some limitations pertaining to the realizability of the cavity parameters in different dielectric media and the existence of spurious responses are discussed.

Design Formula

The design procedure for direct-coupled cavity-type strip-line filters can be evolved from well-known synthesis procedures, [8, 13] currently applied in the design of lumped parameter circuitry. The purpose for such an evolution of a microwave synthesis procedure is two-fold: (1) electrical engineers are, in general, more familiar with circuit theory than with microwave theory, and (2) the effect of each microwave component on the over-all filter performance may be more clearly understood when related to an equivalent circuit parameter.

Bandpass filters can be designed from low-pass prototypes through a low-pass-to-band-pass transformation. The type and complexity of the low-pass prototype is dependent upon the desired frequency response characteristics (skirt selectivity and ripple tolerances in the pass-band). For example, the "maximally flat" response function (sometimes called the Butterworth function) is widely used in filter and i-f amplifier designs because of its mathematical simplicity and monotonic characteristic in the pass-band. Other response functions, such as the Tchebycheff function, also have special applications.

A typical bandpass filter resulting from a low-pass-to-band-pass transformation is shown in Fig. 1A. As will be seen later, this configuration is not as easily realized in strip lines as is the circuit shown in Fig. 1B. It should be noted that the transfer function of the former network has an equal zero distribution at zero and infinite frequencies in the complex frequency plane, whereas the filter shown in Fig. 1B has an unequal distribution with 2n-1 zeros at zero frequency and one at infinite frequency (n is here defined as the number of resonant elements). For this reason, there exists no one-to-one relationship between the response characteristics of these two circuits. However, if the driving-point impedances of the two networks are equated and if the circuit components of one network are expressed in terms of the other, the pole distributions of each network will be identical. Although the zero distributions of the filters differ, the response characteristics are essentially identical in the neighborhood of the pass-band for small filter bandwidths. For large bandwidths, the frequency response characteristic of the filter having an unequal zero distribution will be asymmetrical.

Using transmission line theory, the normalized coupling reactances, $\tilde{X}_m$, and cavity lengths, $\ell_{m,m+1}$, of a direct-coupled cavity-type filter, having a configuration shown in Fig. 2, can be expressed in terms of the lumped parameters of the capacitive-coupled filter [13, 15] (cf., Fig. 1B). The general design equations for an n-stage "maximally flat" bandpass filter are tabulated below.

The normalized reactance of the end discontinuities is

$$\tilde{X}_1 = \tilde{X}_{n+1} \simeq \sqrt{\frac{4}{\pi}\ \frac{f_0}{\Delta f}\ \sin\frac{\pi}{2n}} \qquad (1)$$

and the normalized reactances of all other discontinuities are

$$X_m \simeq \frac{2\sqrt{2}}{\pi}\frac{f_o}{\Delta f}\sqrt{\cos\frac{\pi}{n} - \cos\frac{2(m-1)\pi}{n}} \quad (2)$$

where f_o = center frequency of filter
Δf = filter bandwidth
m = serial number of discontinuity= 2, 3, ...

The separation of the reactance elements is

$$l_{m,m+1} = \frac{\lambda_a}{2\pi\sqrt{\epsilon}}\left[K\pi + \frac{1}{2}\left(\tan^{-1}\frac{2}{X_m} + \tan^{-1}\frac{2}{X_{m+1}}\right)\right] \quad (3)$$

where λ_a = electrical wavelength in air
K = a positive integer
ϵ = dielectric constant of transmission medium

Physical Realization

The most difficult problem in the development of strip line filters is their physical realization in the proposed type of transmission line. Prior to fabrication, it is necessary to (1) choose a suitable dielectric medium, (2) develop a broadband coaxial-to-strip-line transition, and (3) compile a reference library of normalized reactances vs. capacitance-gap spacings for strip line.

For filters requiring high loaded Q-factors and a small insertion loss in the microwave spectrum, the use of an air-dielectric line similar to that developed by AIL [5, 6, 12] is recommended; where physical size is an important consideration at lower frequencies and loaded Q-factors are not large, a high dielectric low-loss material is desired. A study is being made at Melpar to establish the availability of dielectric materials having a high dielectric constant, low loss-tangent, a satisfactory degree of homogeneity, properties suitable for copper-cladding and good environmental characteristics. Although not complete, this evaluation program indicates there are very few high dielectric materials available in a suitable copper-clad laminate form.

Since space reduction is of particular importance in many applications, longer filters resulting from the use of materials having lower dielectric constants can be physically accommodated in a relatively small space by forming the transmission line into a spiral or "snake-like" configuration. Using a very low-loss dielectric such as cementable teflon, a further reduction in size can be obtained over a similar air-dielectric filter.

Having chosen a dielectric medium, a suitable strip-line-to-coax transition is required. To preserve a given filter response, the transition should have a low VSWR over the entire operational frequency spectrum. In Fig. 3A, a transition, employing a UG-290/U BNC connector mounted axially to a strip line having a ground-plane conductor spacing of 1/8 inch, had a VSWR of less than 2:1 from 2 to 3 Kmc. A normal type transition, similar to those used with most Microstrip lines [10] (Fig. 3B), had a narrow frequency response. It should be noted that objects near the edge of this strip line affect the performance of the transition. Asymmetrically located elements in the transverse plane of the "sandwich" type strip lines, such as the extended center conductor of the coaxial connector in the transition of Fig. 3B, cause a perturbation of the electromagnetic field resulting in a marked tendency to radiate. The first satisfactory design of a broad-band, coax-to-strip-line transition built at Melpar in an air-dielectric structure is shown in Fig. 3C. Similarly, a second broadband transition was designed for a strip line using a teflon impregnated fibre-glass dielectric medium. Using a cascaded pair of either of these transitions, the maximum VSWR over a frequency range from 1 to 5 kmc was less than 1.30, as shown in Fig. 4. It should be noted that the terminating load in these tests had a VSWR equal to or less than 1.15 over this spectrum.

The final preparatory step in the realization of a capacitive-coupled strip-line filter is the compilation of normalized reactance data. A series capacitance is easily realized in a strip transmission line by making a gap in the center conductor (Fig. 5A). However, except for very close spacings, it is difficult to calculate the equivalent capacitance of a gap because of the configuration of the fringing field. A reference library of normalized capacitive reactances as a function of gap spacings can be compiled experimentally by making simple tests on one-stage strip-line filters (cf., Fig. 2 for n = 1) having different gap spacings at a specified frequency. For example, having measured the loaded-Q of a one-stage filter, the capacitive reactance of the gap spacing can be computed from Eq. (1). In order to realize the required higher capacitances at the lower frequencies without making the gap spacings unrealistically small (<0.010"), a "dove-tailed" gap shown in Fig. 5B was employed. Typical reference libraries of standard and dove-tailed gaps compiled at 3000 mc in an air-filled line are shown in Fig. 6. As an interesting by-product of these measurements, the dielectric constant of the medium, for other than very small values of Q, can be computed very accurately using Eq (3).

Experimental Results

A six-stage maximally-flat filter, having a ten percent bandwidth, minimum insertion loss, and 40 db rejection at twelve percent from the center frequency, was designed at 3350 mc using the formulas and reference library data discussed above; a photograph of the resulting air-dielectric filter is shown in Fig. 7. It should be noted that dove-tailed gaps are used in the outermost cavities to give the required reactance values. The theoretical and experimental response characteristics of this six-stage filter are shown in Fig. 8.

Filters designed at higher frequencies or having higher loaded-Q factors than the unit shown in Fig. 7 require smaller coupling capacitors, thus eliminating the need for dove-tailed gaps. At lower frequencies, the filters become unrealistically long. To shorten these units, the strip line was photo-etched in a "snake-like" configuration with the reactance gaps located in a linear portion of the line. A filter designed at 715 mc using a teflon-impregnated fiberglas dielectric is shown in Fig. 9. The effect of reflections coming from the impedance discontinuities at the bends is negligible. Theoretical and experimental response characteristics of this unit are shown in Fig. 10.

In many applications, filters must have a large insertion loss over a wide frequency spectrum outside of their fundamental pass-bands. In order to examine spurious responses in strip-line filters, response characteristics of a typical unit centered at 3000 mc were measured from 2500 to 10,000 mc (Fig. 11A). The spurious responses (1) were centered at approximately integer multiples of the fundamental, (2) had about the same Q-factor, and (3) became less discrete as the frequency increased. Spurious responses of this type can be removed with a low-pass filter having a cut-off frequency somewhat less than that of the first spurious response. For example, a low-pass strip-line filter [9] was built with a cutoff frequency of 4000 mc; when cascaded with the above bandpass filter, the composite response characteristics of both filters had no spurious responses up to 10 kmc, as seen in Fig. 11B. Above 10 kmc, it is possible that TE or TM modes of propagation may exist; this region has not yet been investigated.

Some Limitations in Strip-Line Filter Realization

In designing and building direct-coupled strip-line filters using the above techniques, certain limitations exist. The accuracy of the above design formula is a function of the filter requirements. For filters having a loaded Q-factor of ten, it has been empirically observed that the resulting center frequency and loaded Q lie within a $\pm$ 1 percent and $\pm$ 4 percent scatter, respectively. This deviation also includes such errors as those introduced by inaccuracies in the capacitive-reactance reference library and the photo-etching facilities. It has been estimated that photo-etching tolerances of $\pm$ 0.002 inch should be adequate except where exacting requirements dictate a center frequency location of less than $\pm$ 0.5 percent. It is virtually impossible to determine the tolerance requirements since there exist many inseparable variables, each of which is dependent upon the techniques used and the filter requirements.

Where some tolerances cannot be maintained, an alternate way of realizing filters having "exact" center-frequency requirements is to make some provision for tuning. Fig. 12A shows a longitudinal view of a one-stage strip-line filter with capacitive-tuning screws located in both ground-planes. The symmetrical location of these screws is required to minimize radiation that might otherwise exist. In a one-stage filter, a ten percent shift in the center frequency was realized with tuning screws located approximately 0.010" from the center conductor. However, since the loaded Q of the filter was appreciably changed in tuning, as shown in Fig. 12B, the allowable tuning range is, therefore, restricted in certain applications. For a multistage filter, alignment by a method similar to that suggested by Ragan[13] or Dishal [4] is recommended. In this method, all stages are completely detuned (capacitive tuning screws touch the center conductor of the strip line); the quarter-wave shift in the standing-wave pattern at the input circuit is monitored as each stage is progressively tuned.

Conclusions

The design formula and the realization techniques presented here provide the engineer with a formalized design procedure for the development of bandpass filters in a strip line structure. These units have certain advantages over their waveguide and coaxial line counterparts in the u-h-f spectrum. The development of more suitable dielectric-media and the improvement of existing photo-etching techniques will further reduce the difficulties arising in the realization of stringent physical and electrical requirements in strip-line filters.

References

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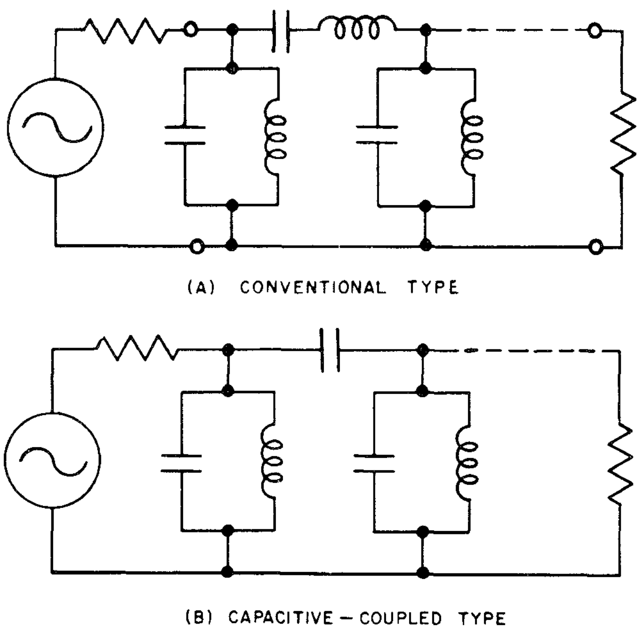


Fig. 1 - Lumped-element bandpass filters.

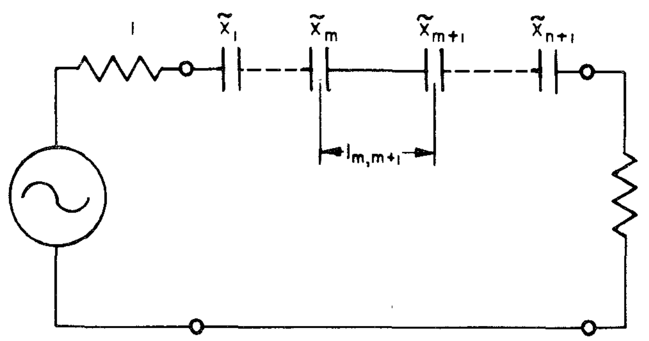


Fig. 2 - Direct-coupled, cavity-type filter.

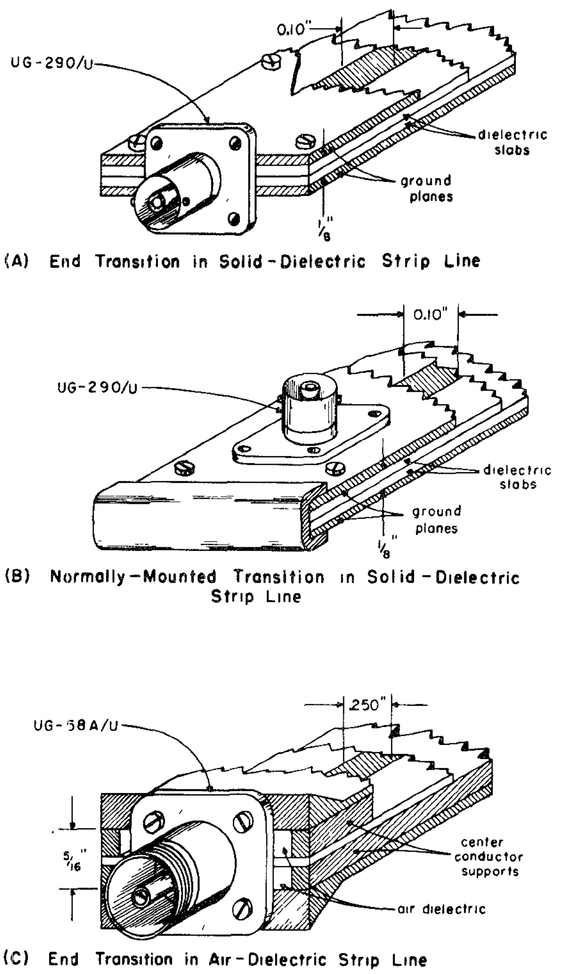


Fig. 3 - Coaxial-to-strip line transitions.

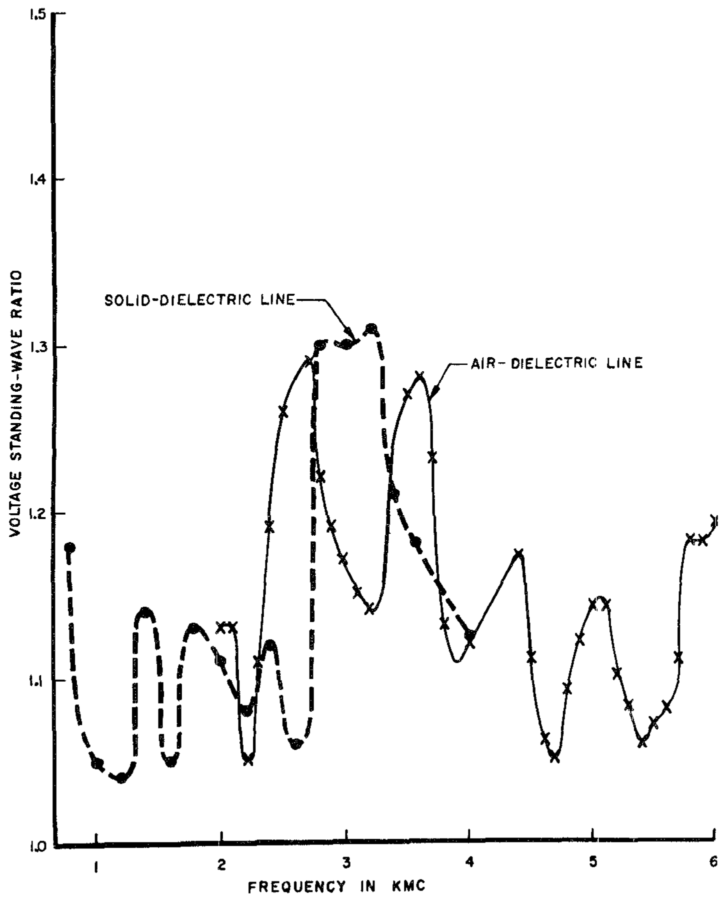


Fig. 4 - Impedance match of coax-to-strip-line transitions.

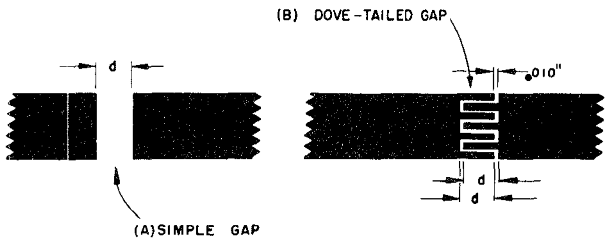


Fig. 5 - Series capacitance realization in strip line.

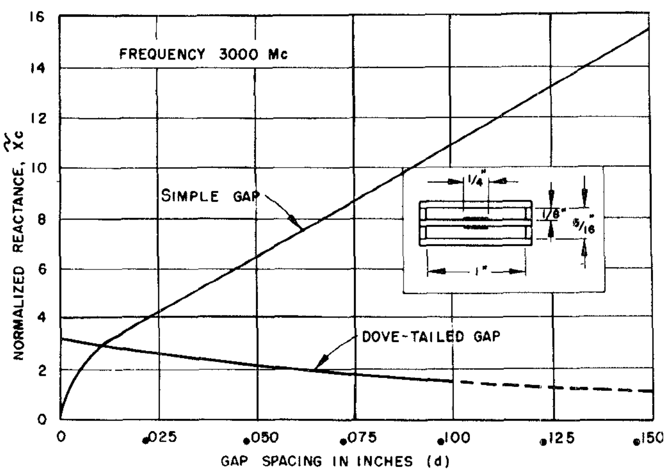


Fig. 6 - Design data for strip-line filters.

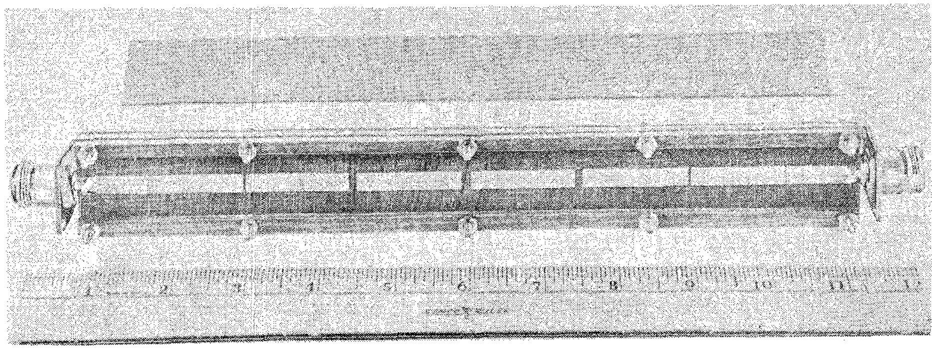


Fig. 7 - Six-stage, bandpass filter in air-dielectric strip line. (fo = 3350 mc)

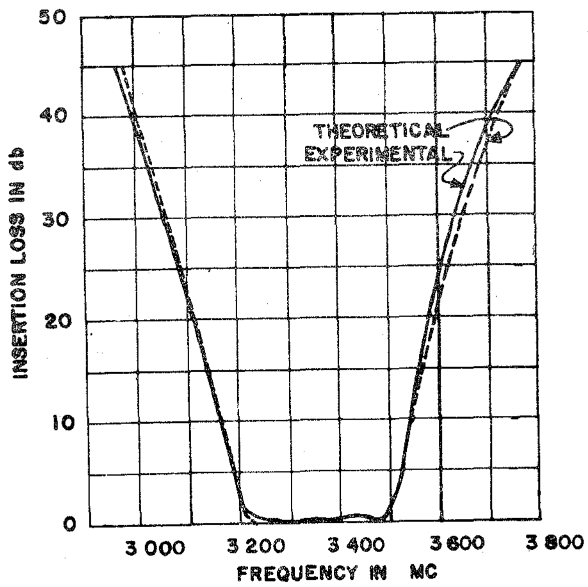


Fig. 8 - Response characteristic of six-stage, strip-line filter with air-dielectric.

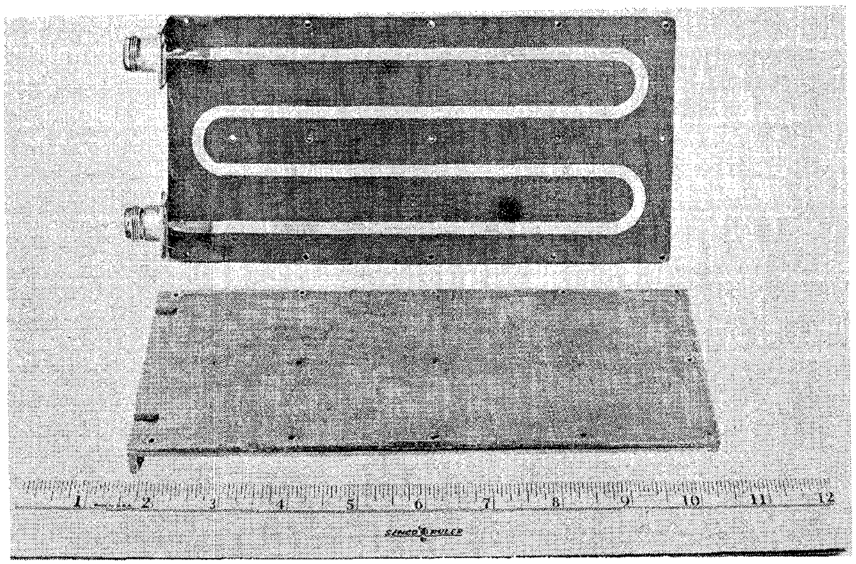


Fig. 9 - Six-stage, bandpass filter in a solid-dielectric strip line. (fo = 715 mc)

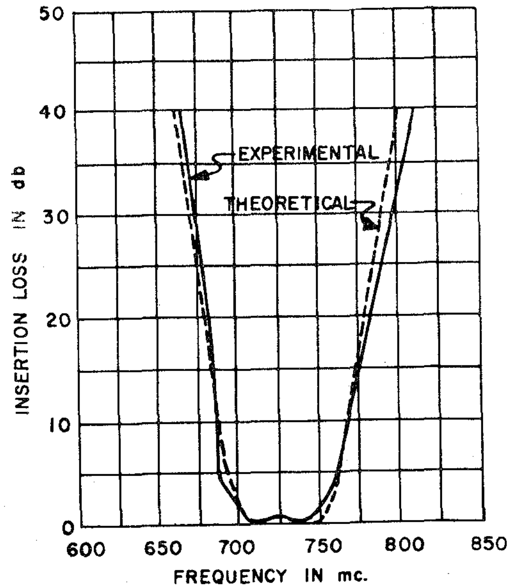
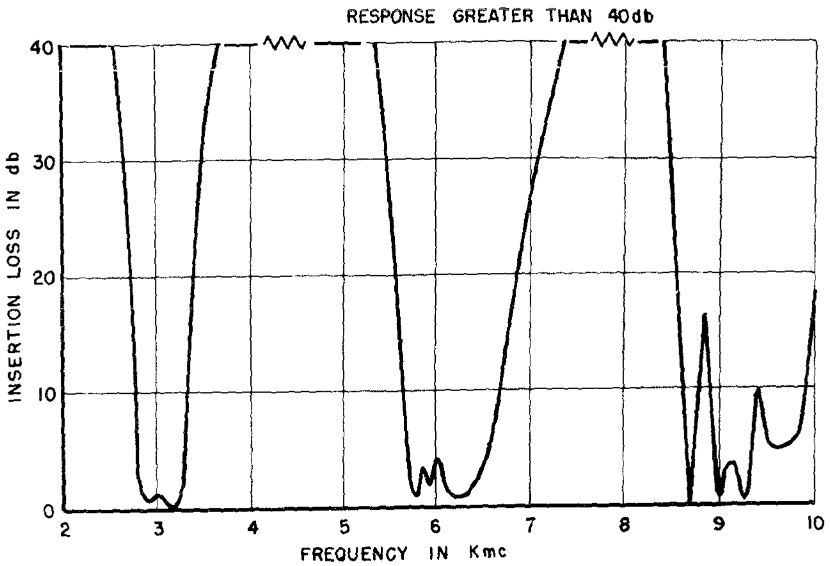
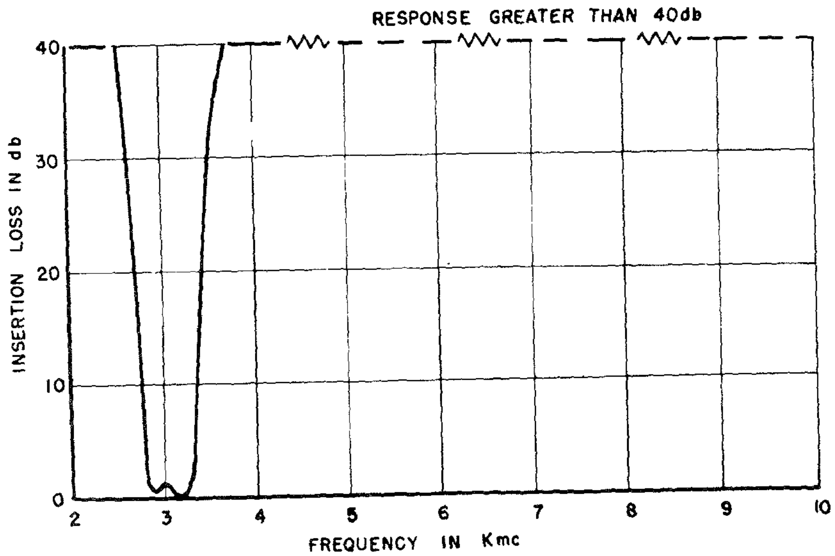


Fig. 10 - Response characteristics of six-stage, strip-line filter in a solid dielectric.

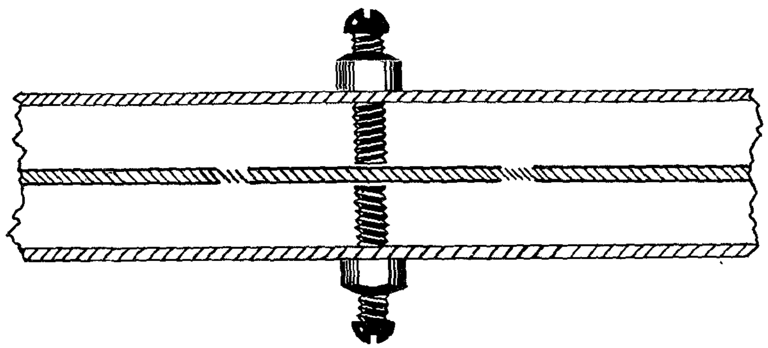


A. FREQUENCY RESPONSE OF A BAND-PASS FILTER

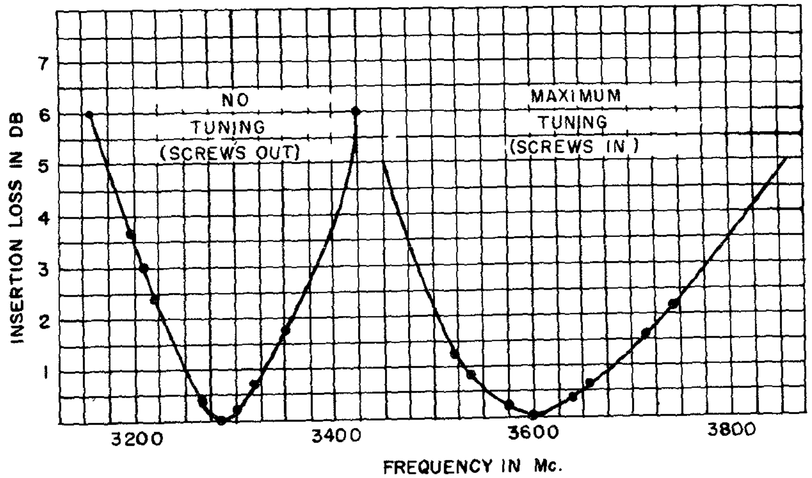


B FREQUENCY RESPONSE OF A CASCADED, BAND-PASS-LOW-PASS FILTER

Fig. 11 - Suppression of spurious responses in a strip-line, band-pass filter.



(A) - CAPACITIVE TUNING SCREWS IN AIR-DIELECTRIC STRIP LINE.



(B) - VARIATION OF RESPONSE CHARACTERISTICS WITH TUNING.

Fig. 12 - Effect of tuning a one-stage, bandpass, strip-line filter.