Capacitively-Coupled Stub Filter

A method of designing an easily realized narrow-band band-stop filter by modifying the exact design procedure [1]–[3] is described herein. Although this design method is not exact, it yields a response curve that follows the desired response over a broad band. The modification consists of replacing each open-circuited high-impedance shunt stub with a short-circuited capacitively coupled stub [4], [5] of medium impedance, i.e., the same order of magnitude as the connecting-line impedance, and slightly less than $\lambda_0/4$ in length, as shown in Fig. 1. The stubs are thus easily realized in coaxial line or strip-line, whereas by the exact method they would (in the case of very narrow-band filters) be difficult to realize unless coupled-line sections were used [1]. A modified three-stub filter having ideally a five per cent stop-band width and 0.1-dB ripple in the pass-band was analyzed. The response in the frequency range 0 to $1.5f_0$ was found to deviate only a negligible amount from the exact response, while the filter was usable over a still broader range. This is clearly illustrated in Fig. 2 which shows the computed response of the three-stub exact-design filter (solid lines) and the computed response of the capacitively-coupled stub filter (dashed lines). In Fig. 2(b), the attenuation loss in the first stop band near $\omega/\omega_0=1$ for the exact-design case is similar to the curve for the capacitively-coupled-stub filter; therefore only a solid line is shown in that region. (The computations for this capactively-coupled stub filter did not extend to the second stop band.) Figure 2(a) shows the response over a smaller range in more detail. Figure 3(a) shows the schematic of the exact-design filter and gives the characteristic impedance of each stub and connecting line, and the terminating impedances, and Fig. 3(b) shows the derived capacitively-coupled stub filter. The high impedance stubs in Fig. 3(a) ($Z_0>1000$ ohms) are impractical for strip-line, whereas the stubs of Fig. 3(b) can be readily constructed.

The formulas for the modification of the stubs are based on the work of Young, et al. [4], and the modification of the exact design yields a circuit that is similar to their narrow-band band-stop filter. These formulas are given in Fig. 1. The calculations of a design are shortened if use is made of a tabulation of the function $F(\phi)$ given in Young, et al. [4]. In this procedure the impedances of the stubs Z_i' are arbitrary, and the extent to which the response of the modified filter will deviate from the exact response depends on the choice of the stub impedances. If the stub impedances are chosen to be as low as possible, the filter response near the first stop band will be close to the exact response while the second stop band will be much wider and will be considerably shifted in frequency from its normal value of three times the frequency of the first stop band. On the other hand, too high stub impedances will preserve the position of the second stop band at $3f_0$, but will result in impractically small capacitive gaps (high capacitance).

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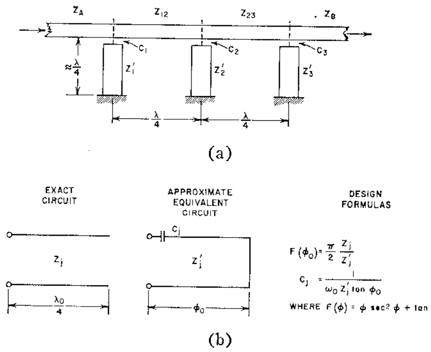


Fig. 1. Strip-line structure for band-stop filters with narrow stop bands.

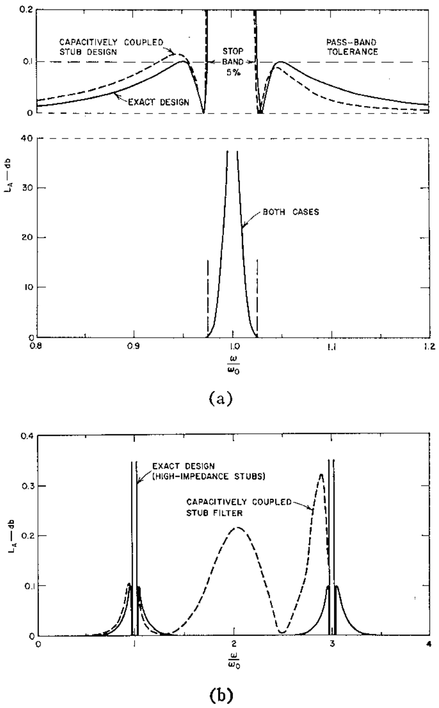
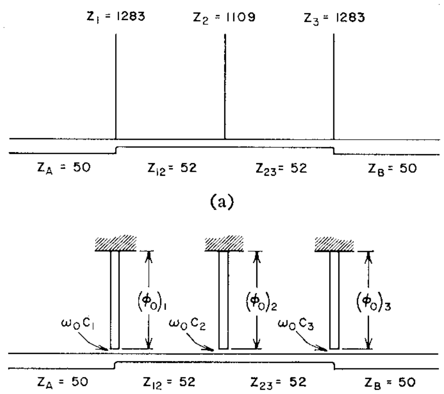


Fig. 2. Computed response of the filter in Fig. 1. (a) Showing details near the first stop band. (b) Showing the first two stop bands.



ALL STUBS HAVE $Z_j' = 151.5$ ohms

$\omega_0 C_1 = \omega_0 C_3 = 0.00239$ mho, $\omega_0 C_2 = 0.00259$ mho

$(\phi_0)_1 = (\phi_0)_3 = 70.1°$, $(\phi_0)_2 = 68.7°$

(b)

Fig. 3. Narrow-stop-band filters. (a) Exact design. (b) Practical realization.

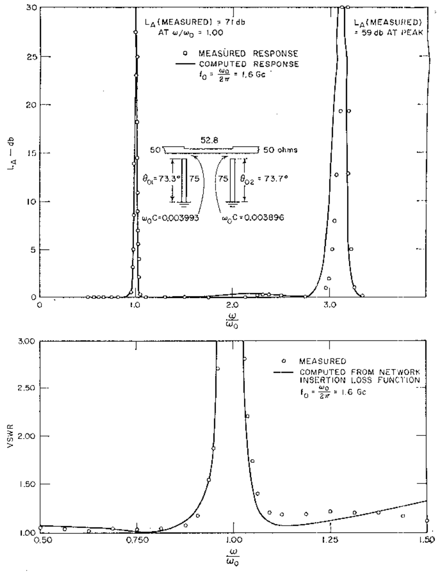


Fig. 4. Schematic diagram of capacitively-coupled stub filter, and its measured and computed responses.

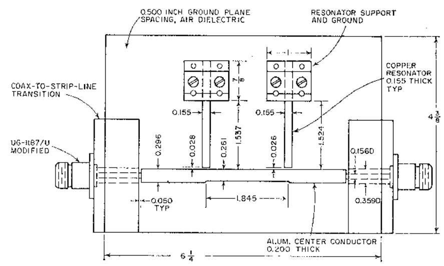


Fig. 5. Sketch of capacitively-coupled stub narrow-band filter with cover plate removed.

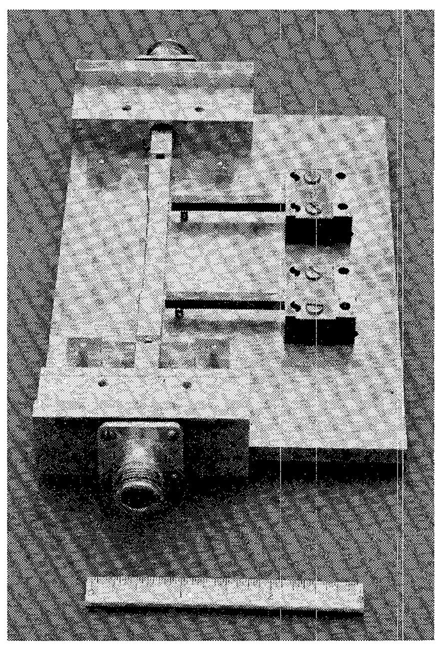


Fig. 6. Photograph of capacitively-coupled stub filter with cover plate removed.

A two-stub filter derived from an exact maximally flat design was constructed in strip-line [5]. The stop-band width is five per cent of the stop-band center frequency $f_0 = 1.6$ Gc/s, and the basic filter has a maximally-flat response. The element values of the low-pass prototype are $g_0 = g_3 = 1$, and $g_1 = g_2 = 1.414$. The exact strip-line filter has stub impedances $Z_1 = 950$ ohms, and $Z_2 = 900$ ohms, and a connecting-line impedance $Z_{12} = 53$ ohms, and 50-ohm terminations. The modified filter is shown in Fig. 4(a), which also shows its computed and measured attenuation loss. Figure 4(b) shows the measured and computed VSWR for a small region near the first stop band. A sketch of the filter is given in Fig. 5, and a photograph is shown in Fig. 6, the cover plate having been removed in both cases. Tuning screws and stub supports with provision for adjusting the length of each resonator permit the coupling gaps and the resonant lengths to be independently adjusted. In this two-resonator design, the second stop band is seen to be considerably wider than the first stop band, unlike the exact design on which it is based. Nevertheless, this design method appears to be useful and capable of yielding accurate designs of practical narrow-band microwave band-stop filters.

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Coupled Rods Between Ground Planes

In a recent paper, Cristal [1] provided graphs of self- and mutual-capacitances (C_g/ϵ and C_M/ϵ) of parallel round conductors between two ground planes as a function of rod separation distance s/b, with rod diameter d/b as a parameter. To realize a distributed element filter, the self- and mutual-capacitance values are first calculated [2], [3]. Then the separation distances and rod diameters are found by using Cristal's graphs which are based on the assumption that the charge on the rod is equally distributed between facing ground planes and adjacent rods.

The subject of this correspondence is to show that after calculating the capacitance values as above, Honey's approximations [4] can be used to determine analytically the geometry of coupled-rod structures with favorable results.

Honey's approximations are:

$$Z_{oe} - Z_{oo} = \frac{120}{\sqrt{\epsilon r}} \ln \coth \frac{\pi s}{2b}$$

$$Z_{oe} + Z_{oo} = \frac{120}{\sqrt{\epsilon r}} \ln \coth \frac{\pi d}{4b}$$

The first filter using Honey's approximation was designed and built in 1962 by E. Cota of these laboratories. Since then various filters in round bar configuration have been realized, ranging from 0.3 per cent bandwidth comb line structure to 72 per cent interdigital types.

The transformation of Matthaei's 10 per cent interdigital filter into circular rods between ground planes by using Honey's approximations is carried out in the Appendix. The transformation calculations, together with Matthaei's filter synthesis procedures, are easily programmed on a computer, thereby eliminating tedious manual calculations.

A comparison of Cristal's procedure and this analytic method is made in Table I.

From the table it is readily seen that the separation distances $s_{k,k+1}/b$ agree better than the rod diameters d_k/b. This method yields slightly higher even and odd mode impedances than Cristal's procedure. Since in this type of geometry, the optimum Q corresponds to a broad range of impedance values, the filter performances using either one of the above realizations are essentially equivalent. Obviously, a round bar configuration will yield a higher Q and is easier to fabricate as pointed out by Cristal.

The end resonators by this method are found to be bigger (comparative to Getsinger's bars) and their respective separation distance to the adjacent resonator smaller than Cristal's procedure. However, Cristal found that the separation distance between the end resonators required empirical adjustment to a value of 0.625 inch closely agreeing with the 0.623 inch value computed with the aid of Honey's approximation.

Thus the described method can be manually calculated or programmed on a computer to yield directly rod diameter and separation distance values that require no graphical interpolation or empirical adjustment.

APPENDIX

The following equations are given:

$$Z_{oe}^k - Z_{oo}^k = \frac{120}{\sqrt{\epsilon r}} \ln \coth \frac{\pi}{2} \frac{s_{k,k+1}}{b} \tag{1}$$

$$Z_{oe}^k + Z_{oo}^k = \frac{120}{\sqrt{\epsilon r}} \ln \coth \frac{\pi}{4} \frac{d_k}{b} \tag{2}$$

$$Z_{oo}^k = \frac{C_{k+1}}{vF_{k,k+1}} \tag{3}$$

$$Z_{oe}^k = \frac{C_{k+1} + 2C_{k,k+1}}{vF_{k,k+1}} = Z_{oo}^k + \frac{2C_{k,k+1}}{vF_{k,k+1}} \tag{4}$$

where,

$$F_{k,k+1} = C_k C_{k+1} + C_{k,k+1}(C_k + C_{k+1}) \tag{5}$$

v = velocity of light in medium of propagation.

From (1) to (5) we get:

$$Z_{oe}^k - Z_{oo}^k = \frac{2C_{k,k+1}}{vF_{k,k+1}} = \frac{120}{\sqrt{\epsilon r}} \ln \coth \frac{\pi}{2} \frac{s_{k,k+1}}{b} \tag{6}$$

$$\frac{60}{\sqrt{\epsilon r}} \ln \coth \frac{\pi}{2} \frac{s_{k,k+1}}{b} = \frac{C_{k,k+1}}{vF_{k,k+1}} \tag{6'}$$

$$Z_{oe}^k + Z_{oo}^k = 2\left(\frac{C_{k+1} + C_{k,k+1}}{vF_{k,k+1}}\right) \tag{7}$$

$$\frac{60}{\sqrt{\epsilon r}} \ln \coth \frac{\pi}{4} \frac{d_k}{b} = \overline{Z_{oo}^k} + \overline{\frac{C_{k,k+1}}{vF_{k,k+1}}} \tag{7'}$$

The impedance-capacitance relationships (3) and (4) define $N-1$ equalities for a set of N elements; thus the separation distances $s_{k,k+1}/b$ are uniquely determined from (6′). In the case of rod diameters, we resort to an averaging process to get N equalities from $N-1$ relationships (7′). The average impedance of a circular rod in this case is computed from the average of the self and mutual capacitances looking to the right and left from the respective circular rod. Thus we rewrite (7′) into the following form:

$$\frac{60}{\sqrt{\epsilon r}} \ln \coth \frac{\pi}{4} \frac{d_k}{b} = Z_{oo}^k + \frac{C_{k,k+1}}{vF_{k,k+1}} \tag{8}$$

which yields N equations for N elements, as required.

In the case of the end resonators, we include the fringing capacitance to one side. From Fig. 1, using (3), we can write:

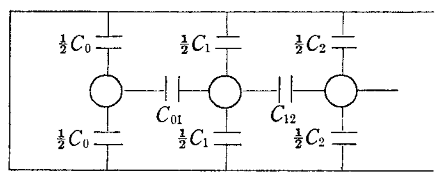

Fig. 1.

Z_{oo}^0 left $= 0$, provided the filter housing end plate is sufficiently far removed,

$$Z_{oo}^0 \text{ right} = \frac{C_1}{vF_{01}} = \frac{1}{v} \frac{C_1}{C_0 C_1 + C_{01}(C_0 + C_1)}$$

Thus

$$\overline{Z_{oo}^0} = 1/2(Z_{oo}^0 \text{left} + Z_{oo}^0 \text{right}) = 1/2(Z_{oo}^0) \tag{9}$$

The remaining $\overline{Z_{oo}^k}$'s, that is from Z_{oo}^1 to Z_{oo}^{k-1} are

$$\overline{Z_{oo}^k} = 1/2(Z_{oo}^k \text{left} + Z_{oo}^k \text{right}) \tag{10}$$

From Matthaei's distributed element filter synthesis, we obtain C_k/ϵ and $C_{k,k+1}/\epsilon$ instead of C_k, and $C_{k,k+1}$, thus (3) and (4) become:

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