Tunable Passive Multicouplers Employing Minimum-Loss Filters*

J. F. CLINE† AND B. M. SCHIFFMAN‡

Summary—**Several multicoupler techniques are described for operating twenty or more UHF transmitters and receivers simultaneously with a single localized antenna system. The types of multicouplers considered include the simple parallel-connected filter type and several distributed-line types, in which the individual branches are separately tuned. The filters used are of the symmetrical, narrow-band, direct-coupled resonator type, designed to obtain minimum center-frequency insertion loss for a given insertion loss in the adjacent channels. Design formulas are given for these filters, and characteristic response shapes are presented. The extra-channel susceptance, which is the principal factor limiting the number of channels obtainable in a single multicoupler, is discussed in terms of the input coupling coefficient, the resonator parameters, and the lengths of the connecting lines.**

INTRODUCTION

THE NUMBER of antennas that can be accommodated in a given air, sea or ground-based installation is limited by considerations of space restrictions, mutual disturbance of radiation patterns, and intercoupling of signals. In certain complex electronic systems now in existence and in others being planned, the large number of individual transmitters and receivers makes it impossible to satisfy the system requirements by a multiplicity of antennas alone, and, therefore, special multicoupler networks involving filters and other components are necessary to permit the sharing of antennas by groups of receivers and transmitters.

This paper has evolved from a study program covering several aspects of multicoupler design. The multicouplers considered here consist of minimum-loss narrow-band filters connected together in different ways. The basic assumptions are that the number of branches is large, *i.e.*, twenty or more, while the pass band of each branch is small in comparison with the minimum channel separation. A typical application is found in the 225 to 400 mc band, with a minimum channel spacing of 2 mc, an adjacent channel isolation of 60 db, a channel-center loss of 1 db or less, and a pass bandwidth sufficient for typical AM, SSB, or NBFM speech modulated signals.

The problems considered here include the design of the minimum-loss filters, the manner of their interconnection, and their effects upon each other from the standpoint of the impedance matching problem. Since these mutual effects involve both the interconnection scheme and the filter design, these two topics must be considered simultaneously. Fixed-tuned, complementary-filter multicouplers are not considered here; instead, the emphasis is on the type of multicoupler in which separate narrow-band channels are independently tunable. Since it may be necessary to change the center frequencies of the individual channels frequently, the preferable method for avoiding excessive impedance mismatch in the multicoupler is to design each branch in such a way as to introduce a minimum of mismatch, rather than to use a matching scheme that involves mutual adjustments of two or more channels at once.

MULTICOUPLER NETWORK CONNECTIONS

In all of the figures, the letters A, LS, R, and T are used to represent antennas, line-stretchers, receivers, and transmitters, respectively. Most of the diagrams are drawn under the assumption that the transmission line sections are of the coaxial or strip-line variety and that the resonant or anti-resonant devices, including those within the filters, are coaxial resonators. However, the waveguide or wire-circuit counterparts may be considered in situations where those techniques are applicable. The results of the filter analysis given in the following section apply equally well to the RF and the microwave regions, since they are expressed in terms of such basic quantities as the loaded and unloaded value of Q, the coupling coefficient, the normalized frequency displacement, and the insertion loss.

Fig. 1 shows two possible multicoupler arrangements. The symbol S represents a two-position switch. Symbols having the same subscript represent parts operating on a common frequency. In the circuit at the left, the isolation between the channels is provided entirely by the filters in the multicoupler, on a frequency difference basis. In the circuit at the right, additional isolation is provided by the two-port antenna system, on the basis of either phase cancellation or field orientation. In general, an antenna system having any number of ports greater than unity can be used in this application. An example of a three-port microwave antenna system of this type is described by Honey and Jones.[1]

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† Stanford Res. Inst., Menlo Park, Calif.
‡ Varian Associates, Inc., Palo Alto, Calif.

[1] R. C. Honey and E. M. T. Jones, "A versatile multi-port biconical antenna," 1957 IRE NATIONAL CONVENTION RECORD, pt. 1, pp. 129–137.

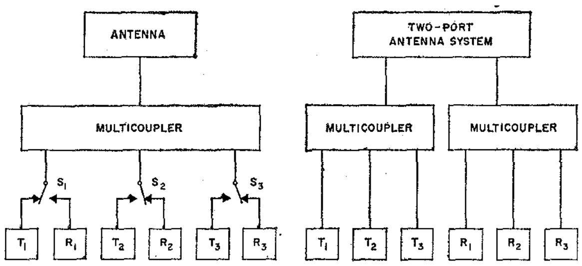

Fig. 1—Two possible antenna-multicoupler systems.

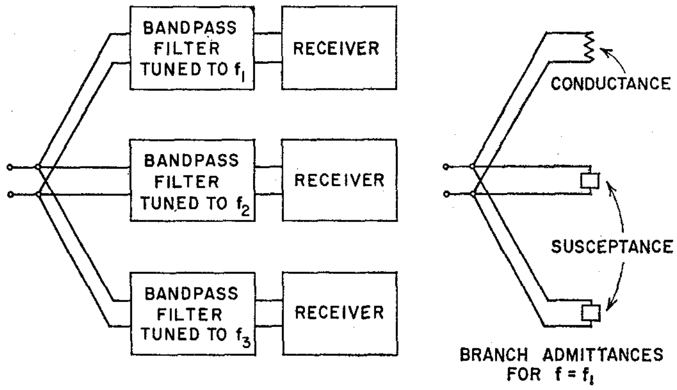

Fig. 2—Parallel-connected multicoupler.

Fig. 2 illustrates a common-junction parallel-connected multicoupler. The admittance, looking from the left into the common junction, is measured at the frequency f_1, which is the center frequency of the upper branch. The input admittance of the upper filter at this frequency is a conductance approximately equal to the characteristic admittance of the connecting lines. The other filters appear as small susceptances, the values of which, in general, are increased considerably by the effects of the short connecting lines between the junction and the filters. The sum of these extra-channel susceptances appearing at the common junction produces an admittance mismatch at this point. This mismatch is the principal factor limiting the number of permissible branches in the multicoupler. It can be reduced by shortening the connecting lines, improving the junction design, and by applying the usual compensation techniques, but it cannot be eliminated entirely over a whole band of frequencies.

Figs. 3 through 6 show several ways of connecting filters in a distributed fashion along a transmission line. In Fig. 3, LS_N is adjusted so that the admittance looking toward the right from the Nth junction is as near zero as possible at the operating frequency of R_N and T_N. At the same frequency each of the other filters looks like a small capacitance, which can be compensated locally. Once LS_N has been adjusted, LS_{N-1} is then adjusted so that the admittance looking toward the right from the $N-1$ junction is as near zero as possible at the operating frequency of R_{N-1} and T_{N-1}, etc.

A possible circuit-packaging arrangement for the different sections in Fig. 3 is suggested in Fig. 4, where

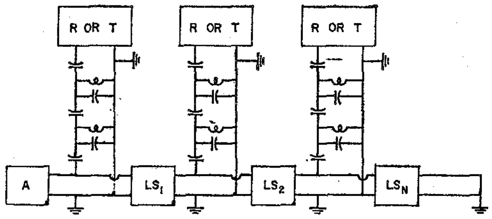

Fig. 3—Distributively-connected multicoupler.

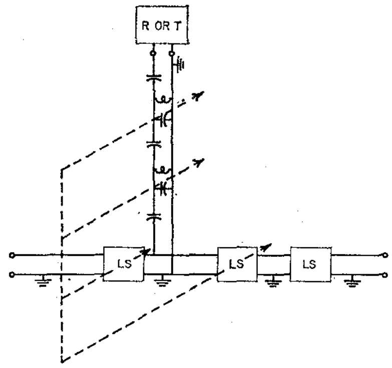

Fig. 4—Single multicoupler section.

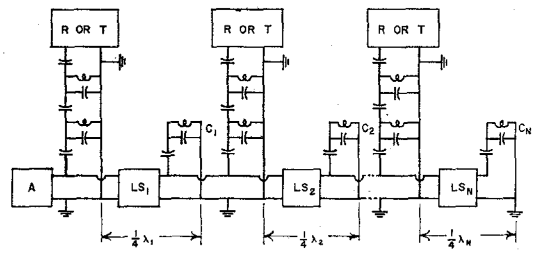

Fig. 5—Line-stretcher and isolation-resonator connections.

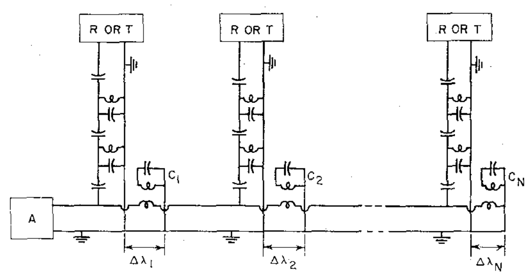

Fig. 6—Circuit employing isolation resonators only.

the entire diagram represents a single section of Fig. 3. Three line stretchers are used instead of one, and two of the line stretchers are gang-tuned together with the band-pass filter. This makes each section nearly independent of the others as far as tuning is concerned. The third line stretcher is adjusted separately, depending upon where the particular section in question is placed in the network chain. Once set, it ordinarily is not reset during tuning operations.[2]

[2] The idea of connecting and ganging filters and line stretchers in a separate package was suggested verbally by L. E. Friedman, formerly with Balco Research Laboratories of Newark, N. J.

If it is permissible to restrict the tuning range of each ection in Fig. 3 so that the center frequencies always ncrease monotonically with section number, the line tretchers can be eliminated and their function can be erformed by means of a mechanical arrangement in vhich the filter ports slide along the transmission line. Even if such a restriction is not permissible, the line tretchers still can be eliminated if a means can be de-vised for sliding the filter ports past each other or plug-ging them in at different points along the line. Another nethod is to use a line that is long enough so that each liding filter port has a freedom of motion of at least a alf wavelength at its lowest frequency of operation.

The circuit of Fig. 5 avoids the interdependence ffects of the adjacent line stretchers by the interposi-ion of resonators coupled to the line. For example, the esonator C_1 is adjusted to place, in effect, a short cir-uit in parallel with the line at the frequency of opera-ion of R_1 and T_1. The line stretcher LS_1 then is adjusted o separate the cavity and the filter by a quarter wave-ength, so that the filter port is, in effect, in parallel vith an open circuit. The new problems in Fig. 5 are ssociated with the design of these added resonators nd their coupling to the line, with the object of pro-lucing an isolating effect in the pass band while produc-ng as little effect as possible in adjacent channels.

Fig. 6 illustrates a scheme for avoiding the use of ither sliding ports or line stretchers. The coupled esonators C_1, C_2, etc., act as open circuits at their cen-er frequencies, and they are placed as close as possible o the filter junctions; that is, the distances $\Delta\lambda$ in the liagram are made as small as possible.

Other applicable types of multicouplers have been lescribed by Carlin,[3] by Cohn and Coale,[4] and by Lewis nd Tillotson.[5] Carlin's multicoupler has the practical lisadvantage that each section operates on the prin-iple of a balanced bridge, in which the antenna imped-nce is one arm of the bridge, so that a careful balance nust be maintained in order to provide the required solation. The other two multicouplers employ direc-ional filters that are constant-resistance networks and re best suited for use in fixed-tuned multicouplers vhere the pass bands are more nearly contiguous. They re more complex than is necessary in the present ap-plication, where the pass bands are narrow in compari-son with the channel separations. Multicouplers em-ploying resistance-terminated lines also have been used, especially for receiver application in strong signal areas where the added losses can be tolerated, but the addi-tional loss makes their use undesirable in cases where a more satisfactory solution can be found.

[3] H. J. Carlin, "UHF multiplexer uses selective couplers," *Electronics*, vol. 28, pp. 152–155; November, 1955.

[4] S. B. Cohn and F. S. Coale, "Directional channel-separation filters," Proc. IRE, vol. 44, pp. 1018–1024; August, 1956.

[5] W. D. Lewis and L. C. Tillotson, "A non-reflecting branching filter for microwaves," *Bell Sys. Tech. J.*, vol. 27, pp. 83–95; January, 1948.

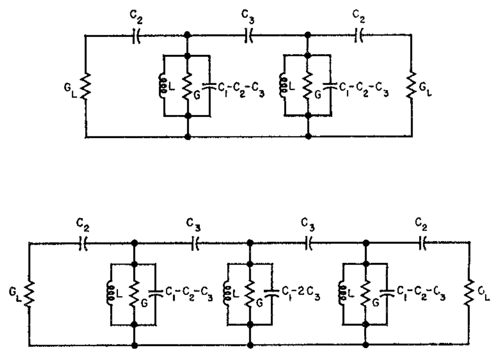

Fig. 7—Equivalent circuits of direct-coupled cavity-resonator filters near resonance.

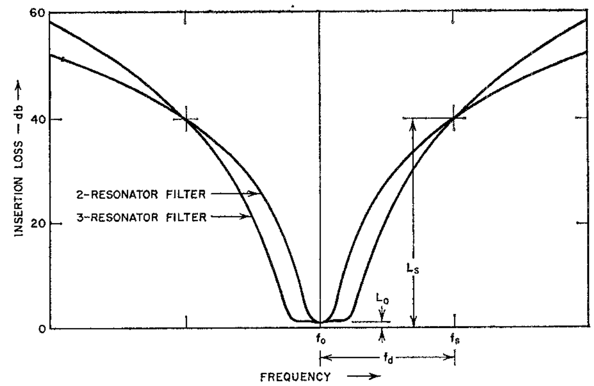

Fig. 8—Notation for filter response curves.

Filter Designs

Design Basis

The narrow-band filters considered here are of the symmetrical, direct-coupled resonator type. Primary attention has been given to two and three-resonator capacitively coupled filters, equivalent circuits for which appear in Fig. 7. The terminal load conductances are denoted as G_L, the terminal coupling capacitances are denoted as C_2, and the internal coupling capacitances are denoted as C_3. The resonators each are represented as a parallel combination of inductance L, conductance G, and a capacitance of which the major part is denoted as C_1. All resonators in any one filter are identical and have a value of unloaded Q that is denoted as Q_U. It may be noted that the total of all capacitances connected to the upper terminal of each resonator is exactly equal to C_1. The value C_1 is a resonator design parameter which is not set by the filter response specifications, but may be fixed at any arbitrary value within the limits imposed by the resonator geometry. Terminal and internal coupling coefficients are defined as $k_T = C_2/C_1$ and $k = C_3/C_1$, respectively.

Fig. 8 shows typical response curves for the two types of filters and introduces certain symbols needed for their description. The center frequency is denoted as f_0 and the insertion loss in decibels at that frequency is denoted as L_0. A design frequency f_S is defined at a point in the filter stop band where the insertion loss in decibels is

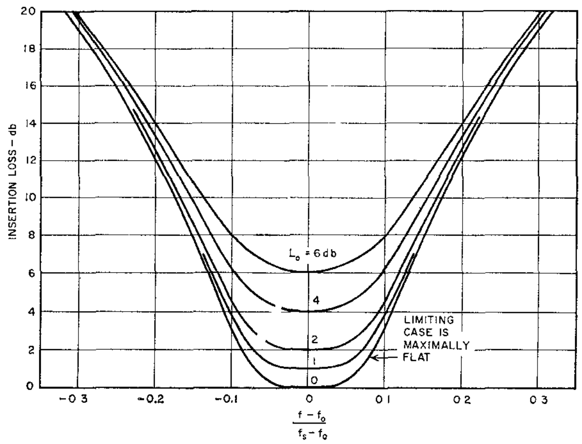

Fig. 9—Two-resonator minimum-loss response curves for $L_s=40$ db.

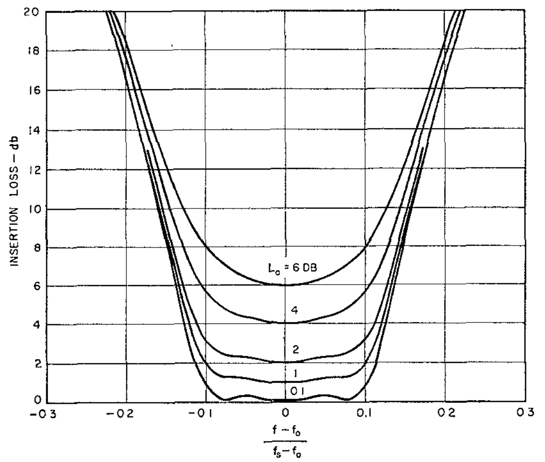

Fig. 10—Three-resonator minimum-loss response curves for $L_s=60$ db

made equal to a design value L_S. The difference between f_s and f_0 is denoted as f_d. The symbol ω, appearing with various subscripts at other points in this paper, is used in the usual way to represent 2π times the value of f carrying the same subscript. The symbols f and ω without subscripts represent the general frequency variable.

The filters described here are symmetrical filters designed on a *minimum loss* basis, *i.e.*, the values of f_0, f_d, and L_S are specified and k_T and k are then adjusted to produce the smallest value of L_0 for a given value of Q_U. Whatever response shapes result from following this design procedure are then accepted if they are adequate for the particular application. Some typical response curves are shown in Figs. 9 and 10.

Two-Resonator Filters

The general expression for insertion loss, in decibels, as derived by Cohn and Shimizu,[6] is

$$L_I = 10 \log_{10} \left\{ \left[\frac{(1+Q_T/Q_U)^2}{2kQ_T} + \frac{kQ_T}{2} \right]^2 + 2\left[\frac{(1+Q_T/Q_U)^2}{k^2} - Q_T^2 \right] u^2 + \frac{4Q_T^2}{k^2} u^4 \right\} \quad (1)$$

where the normalized frequency variable is defined by $u=(f-f_0)/f_0$ and Q_T is the Q of each cavity due to terminal loading only.

The first term of (1) determines the midband loss and is a number that is generally less than 10. The second term, which is significant only for filters with large degrees of overcoupling or undercoupling, has a coefficient which is the difference between two approxinately equal quantities and is, therefore, small. The third term, however, increases as the fourth power of u and is therefore the only significant term in the stop band. Hence the insertion loss in decibels at frequency f_s is given quite accurately by

$$L_s = 10 \log_{10} \left(\frac{4Q_T^2}{k^2} u_s^4 \right) \quad (2)$$

where u_s is the value of u for $f=f_s$, *i.e.*, $u_s=(f_s-f_0)/f_0$. It can now be seen that the ratio Q_T/k must be held constant in the process of finding a minimum-midband loss design for specified values of L_s and u_s.

First, u is set equal to zero in (1). Then the equation is rearranged to give the band-center loss, in decibels, as

$$L_0 = 20 \log_{10} \left[\frac{1}{2} \left(\frac{Q_T}{k} \right) \left(\frac{Q_T+Q_U}{Q_TQ_U} \right)^2 + \frac{Q_T^2}{2} \left(\frac{k}{Q_T} \right) \right]. \quad (3)$$

The argument of the above equation is differentiated with respect to Q_T while holding Q_U and (Q_T/k) invariant. The derivative is then set equal to zero, and after simplifying the result with the aid of (2), the condition for minimizing L_0 is found to be

$$Q_U u_s = \frac{10^{L_s/40} \sqrt[4]{1+Q_T/Q_U}}{\sqrt{2Q_T/Q_U}} . \quad (4)$$

When (2) and (4) are substituted in (3) we obtain the following expression for the minimum center-frequency loss of a symmetrical two-resonator filter, in decibels:

$$L_0 = 10 \log_{10} \left[\frac{1}{4} \left(\frac{Q_T}{Q_U} + 1 \right) \left(\frac{Q_T}{Q_U} + 2 \right)^2 \right]. \quad (5)$$

In practice, the value of Q_T/Q_U which satisfies (5) can be found either by a graphical method, by an iterative process, or by solving the cubic equation. This value of Q_T/Q_U inserted in (4) gives a simple relation between the minimum value of Q_U, the stop band insertion loss L_S, and the normalized frequency variable u_s which will yield that insertion loss.

Finally, the value of the internal coupling coefficient k is obtained by rearranging (2) as follows:

$$k = 2Q_T u_s^2 / 10^{L_s/20}. \quad (6)$$

[6] S. B. Cohn and J. Shimizu, "Strip Transmission Lines and Components," Stanford Res. Inst., Menlo Park, Calif., Project No. 1114, Quart. Prog. Rep. 2, Contract No. DA 36-039SC-63232, pp. 4–6; May, 1955.

Thus the electrical design of the two-resonator filter has been completely specified by (4) through (6) for given values of Q_U, L_s, and u_s.

Three-Resonator Filter

A similar procedure is followed for the three-resonator filter. From expressions derived by Taub and Bogner,[7] the midband insertion loss in decibels is given by

$$L_0 = 20 \log_{10} \frac{1/Q_L^2 Q_U + 2k^2/Q_L}{2k^2/Q_T} \tag{7}$$

where Q_L is defined by $1/Q_L = 1/Q_T + 1/Q_U$. Eq. (7) applies to a symmetrical filter, and our notation, not that of Taub and Bogner, is used. In the case of the optimum three-resonator filter it is impossible to achieve a perfect maximally-flat or Tchebycheff response in a symmetrical filter having internal losses, as pointed out by Dishal and Sellers[8] and by Taub and Bogner.[7] By way of comparison, however, the symmetrical design derived below has the following advantages over the exact response shape design:

1) A better pass-band match is possible.

2) Midband loss is lower for the same values of f_s, L_s, and Q_U. As derived by Taub and Bogner from expressions given by M. Dishal[9] the loss, in decibels, is

$$L_I = L_0 + 10 \log_{10} \left\{ 1 + \left[\frac{b^2 - ca}{c^2} \right] (2u)^2 + \left[\frac{a^2 - 2b}{c^2} \right] (2u)^4 + \frac{(2u)^6}{c^2} \right\} \tag{8}$$

where

$$a = \frac{2}{Q_L} + \frac{1}{Q_U}, \qquad b = 2k^2 + \frac{2}{Q_L Q_U} + \frac{1}{Q_L^2},$$

and

$$c = \frac{2k^2}{Q_L} + \frac{1}{Q_L^2 Q_U}.$$

The insertion loss at the normalized frequency u_s (corresponding to f_s in Fig. 10) as determined from (7) and (8), is given to an excellent approximation by

$$L_s = 20 \log_{10} \left(\frac{4Q_T}{k^2} u_s^3 \right). \tag{9}$$

It is apparent from the above that Q_T/k^2 is the quantity that must be held constant in order to determine the minimum midband loss of a three-resonator filter for a constant value of L_s. The analysis then proceeds in a manner similar to that of the two-resonator case.

The design relation for this case is thus found to be

$$Q_U u_s = \frac{10^{L_s/60} \sqrt[3]{1 + Q_T/Q_U}}{\sqrt[3]{4 Q_T/Q_U}}. \tag{10}$$

By combining (7), (9) and (1) we then obtain an explicit formula for Q_T/Q_U

$$\frac{Q_T}{Q_U} = \frac{3}{2} \left(\sqrt{1 + \frac{8}{9} (10^{L_0/20} - 1)} - 1 \right). \tag{11}$$

The value of the internal coupling coefficient k is obtained by rearranging (9) as follows:

$$k = \frac{2\sqrt{Q_T u_s^3}}{10^{L_s/40}}. \tag{12}$$

Thus the electrical design of a symmetrical three-resonator filter having minimum insertion loss is completely specified by (10) through (12).

Approximate Design Formulas

Certain approximations, accurate to within 5 per cent for values of L_0 less than 1 db, can be obtained for the quantities $Q_U u_s$ and Q_T/Q_U in (4), (5), (10), and (11). Design formulas for the two-resonator filter, based on these approximations, can be written as

$$k \approx 1.41 \frac{\omega_d}{10^{L_s/40} \omega_0} \tag{13}$$

$$k_T \approx 1.19 \frac{\sqrt{G_L \omega_d L}}{10^{L_s/80}} \tag{14}$$

$$Q_U \approx 6.15 \frac{10^{L_s/40} \omega_0}{L_0 \omega_d} \tag{15}$$

while corresponding formulas for the three-resonator filter are

$$k \approx 1.57 \frac{\omega_d}{10^{L_s/60} \omega_0} \tag{16}$$

$$k_T \approx 1.31 \frac{\sqrt{G_L \omega_d L}}{10^{L_s/120}} \tag{17}$$

$$Q_U \approx 8.21 \frac{10^{L_s/60} \omega_0}{L_0 \omega_d}. \tag{18}$$

The value L in (14) and (17) is the value of the resonator equivalent inductance shown in Fig. 7.

Factors Limiting the Number of Branches

Each multicoupler branch in Figs. 2 through 6, at band center, presents to the junction an input admittance approximately equal to G_L. Outside its pass band, each branch presents an input admittance which is essentially a susceptance, described here by the term *extra-channel susceptance.*

[7] J. J. Taub and B. F. Bogner, "Design of three-resonator band pass filters having minimum insertion loss," Proc. IRE, vol. 45, pp. 681–686; May, 1957.

[8] M. Dishal and B. Sellers, "Design of three-resonator dissipative band-pass filters having minimum insertion loss," Proc. IRE, vol. 46, p. 498; February, 1958.

[9] M. Dishal, "Design of dissipative band-pass filters producing desired exact amplitude-frequency characteristics," Proc. IRE, vol. 37, pp. 1050–1069; September, 1949.

The multicoupler is set up in such a way that the center frequency of each branch is different from that of any other branch. In general, the extra-channel susceptance of each branch, measured or calculated at the center frequency of another branch, is small in comparison with G_L. However, the total effect of all of these small extra-channel susceptances is the principal factor limiting the permissible number of branches. Consequently, it is important to make the extra-channel susceptance of each branch as small as possible. Each of these extra-channel susceptances is equal to the extra-channel susceptance of its filter, transformed through the short length of conductor connecting the filter port to the junction or branch point. In the usual case, this is approximately equal to the sum of the extra-channel susceptance of the filter alone plus the lumped capacitance of the short conductor to ground. At VHF and higher frequencies, the magnitudes of the various quantities are such that the capacitance of this short conductor is the larger of these two components unless the conductor is less than a small fraction of an inch in length. This makes it important to design the junction in such a way as to minimize this effect as much as possible.

In the parallel-junction multicoupler shown in Fig. 2, the extra-channel susceptances add directly at the junction. If it is assumed, for simplicity, that each susceptance consists entirely of the lumped susceptance of the short line between the junction and the filter port, the total susceptance in, for example, a 20-channel multicoupler is 20 times the average lumped susceptance of one branch line. The largest total susceptance that can be compensated by practical means over a two-to-one frequency band is probably about equal to G_L. If the characteristic impedance of the lines is also equal to G_L, and if the average frequency is taken as 300 mc, for example, the lines can average no more than about 0.2 inch in length. When the added effects of the filters are considered, this length must be reduced even further. Consequently, the distributed type of multicoupler shown in Fig. 3 seems to offer more promise when there are more than five or ten branches, since the extra-channel susceptance is not concentrated at one point. It appears quite possible to design the distribution line so that the branch susceptances in effect become part of the distributed shunt susceptance of the distribution line itself. In this case, the connecting line effects become less important, and attention can be turned to the extra-channel susceptances of the filters themselves.

It is easily shown that in either the two or the three-resonator filter, the extra-channel susceptance is essentially unchanged when the second resonator is short-circuited. The value of the extra-channel susceptance then is given approximately by

$$B_{xc} \approx \omega k_T C_1 + \omega k_T^2 C_1 \frac{\omega_0^2}{\omega_0^2 - \omega^2} \qquad (19)$$

where ω is 2π times the frequency of measurement or calculation and ω_0 is 2π times the center frequency of the filter in question. The value of k_T is given by either (14) or (17) for the minimum-loss filter designs described in this paper. The minimum-loss filter is well suited to multicoupler applications because of its small value of k_T. For the parameters of other filter designs having any number of resonators, the reader is referred to a recent article by Cohn.[10]

The last term in (19) represents a series combination of inductance and capacitance that is resonant at $\omega = \omega_0$. This series combination, in parallel with C_2, as indicated by the first term, forms the extra-channel equivalent circuit for the filter. This is shown schematically in the upper part of Fig. 11, which represents a group of seven filters tuned to relative frequencies of $\omega_0/\omega = 0.985$, 0.99, etc., as indicated. In the case where $\omega_0/\omega = 1$, the approximation upon which (19) is based is not valid, and the input admittance is approximately equal to G_L. The bar charts below the equivalent circuits illustrate qualitatively the way in which the two branch susceptances values vary with ω_0. The value of ω is the same in all cases. The left hand branch susceptance, ωC_2, increases with decreasing ω_0 because k_T, and hence C_2 also, increases as ω_0 decreases. For values of ω_0 far removed from ω, the susceptance of the right hand branch, represented by the last term in (19), becomes negligible. Also, its sign changes when ω_0 passes through ω. Consequently, when there are a large number of channels spread over a wide frequency range, it may be possible in some cases to neglect the last term in (19) from the standpoint of its average effect, and the average extra-channel susceptance of a filter then reduces to that of C_2 alone.

Although this paper, as stated in the Introduction, has evolved from a study program in which the multicouplers under consideration operate in the 225 to 400 mc band with twenty or more branches, some of the first experimental measurements on equipment designed from the above formulas were made in connection with another multicoupler study program[11] conducted in the HF range and involving only four branches. Figs. 12 and 13 show the measured insertion loss and admittance of a four-branch, parallel-connected multicoupler, in which each branch consists of a minimum-loss filter containing three resonant circuits. In these measurements, the center frequencies were set at 17, 18, 19, and 20 mc. The admittance was measured at the common junction of the four filter ports and the heavy segments of the curve define bandwidths of 0.2 mc. The shunt susceptance of the junction connections at 20 mc, normalized with respect to the characteristic admittance, was measured separately and found to be 0.29.

[10] S. B. Cohn, "Direct-coupled resonator filters," Proc. IRE, vol. 45, pp. 187–196; February, 1957.

[11] Sponsored by U. S. Navy Dept., Bureau of Ships, under contract NObsr-72765.

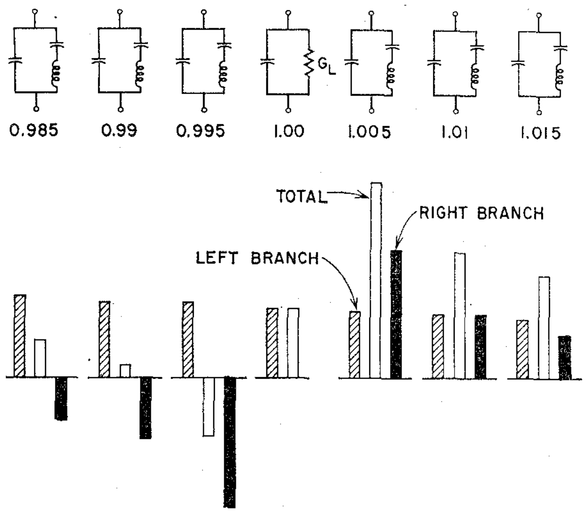

Fig. 11—Multicoupler branch susceptance values.

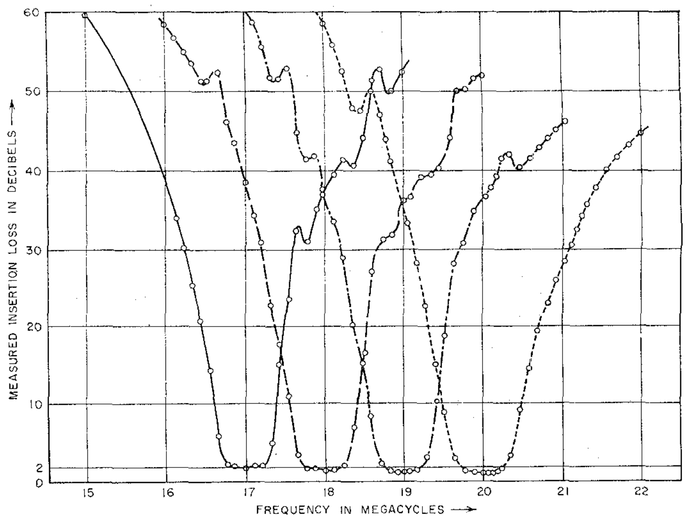

Fig. 12—Measured insertion loss in a four-branch multicoupler.

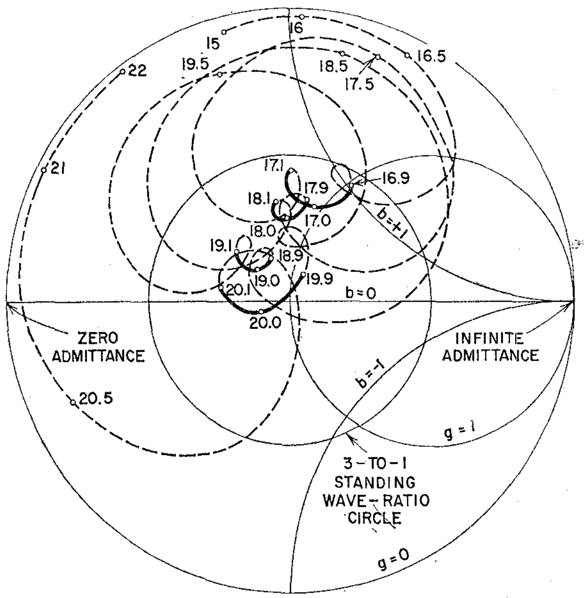

Fig. 13—Measured admittance in a four-branch multicoupler.

The filter centered at 20 mc, tested alone, had values of $L_0=1$ db, $L_s=30$ db, and $\omega_d/\omega_0=0.05$. The filter centered at 17 mc had values of $L_0=1.2$ db, $L_s=37$ db and $\omega_d/\omega_0=0.059$. The resulting values of Q_U given by (18) are 500 and 480, respectively, which agree reasonably well with a value of 530 obtained in separate measurements on the coils alone. The higher loss of about 1.9 db at 17 mc in Fig. 12 is due to the fact that a mismatch exists at that frequency, as shown in Fig. 13. The ripples in the stop band of each branch in Fig. 12 occur in or near the frequencies of the pass bands of the other branches. They can be explained in terms of the shunting action at the junction produced by each filter in and near its pass band. Generally speaking, this is not detrimental to the operation of the multicoupler.

Conclusions

When the pass band of each channel is much narrower than the minimum channel separation, an effective multicoupler may be formed by connecting together several narrow-band filters, either in parallel at a common junction or distributed along a transmission line. The distributed connection is superior to the parallel connection when the number of branches in the multicoupler becomes larger than about five or ten. The directional filter approach to multicoupler design is more complicated than necessary in the present case and is best suited to contiguous pass band applications, which are not considered here.

The narrow-band filters can be designed to produce a minimum center-frequency insertion loss for given values of channel spacing, adjacent-channel insertion loss, and unloaded resonator Q. Symmetrical two and three-resonator filters designed on this basis have response functions that are suitable for many applications, including the present one. In the case of the three-resonator filter, the center-frequency insertion loss is smaller than that obtained in the maximally-flat, unsymmetrical three-resonator filter.

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