Band Pass Filter, Band Elimination Filter and Phase Simulating Network for Carrier Program Systems

By F. S. FARKAS, F. J. HALLENBECK, F. E. STEHLIK*

A paper by Leconte, Penick, Schramm and Wier[1] discusses the system aspects of 8-kc program circuits over carrier facilities and outlines the functions of several filters and networks. This paper describes in detail two of the filters and one network. These are:

1. The channel selecting crystal band pass filter used at program terminals of all broad-band carrier systems,
2. The band elimination filter which blocks the program at branching points on type K carrier systems,
3. The network used at branching points on type K carrier systems to simulate the phase shift of the band elimination filter.

Channel Selecting Crystal Band Pass Filter

An IMPORTANT component of the modulator-demodulator circuit at the carrier program terminal is the band pass filter which selects the lower side band resulting from modulation of the audio frequency program material with the 88-kc carrier. This step of modulation locates the program frequencies in their allotted position in the carrier frequency spectrum of the standard broad-band terminal.

System flexibility requires that long program circuits be established by tandem connections of carrier links. A link consists of a transmitting and a receiving carrier program terminal connected by the appropriate transmission medium. The original objectives were based on a ten-link carrier circuit. This means that each terminal must introduce no more than five per cent of the total allowable system distortion. Assuming the band filter introduces the major part of the terminal distortion it is seen that the requirements placed on each band filter are extremely severe.

One of the transmission objectives of the system is to transmit audio frequencies as low as 50 cps. Hence the band filter must transmit the wanted carrier frequency sideband to within 50 cps of the carrier and must suppress the unwanted sideband beginning at 50 cps above the carrier. This sharp cut-off and the need for low distortion in the pass band requires the use of filter elements with so little dissipation that the only possibility of realizing the desired performance is by the use of quartz crystal elements.

In addition to suppressing the unwanted sideband above the carrier the filter must also provide sufficient discrimination above and below the pass

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band to prevent crosstalk of the adjacent message channels into the program channel.

The necessity of using quartz crystal elements limits the maximum band width of filter which can be realized. This limitation is the result of the comparatively poor electromechanical coupling of quartz.[2] The resulting filter band width is 8.5 kc with the upper cut-off located near the 88-kc carrier. This is slightly greater than the 8-kc nominal band width of the system.

The crystal band pass filter designed for the single sideband program channel weighs approximately 30 lbs. and occupies 7 inches of mounting space on a standard 19 inch relay rack. A total of 44 filter components are required for its construction, half of which are balanced quartz crystal plates. The remaining components consist of eight adjustable air capacitors, three fixed mica capacitors, seven balanced retardation coils, three of which are adjustable, and four resistors. A schematic which shows the relative placement of these parts in the filter is given in Fig. 1.

The measured insertion loss characteristic of the filter between resistive terminations is shown in Fig. 2. The pass band and the vicinity of the upper cut-off are given in greater detail in the enlarged characteristics of Figs. 3 and 4. The extreme sharpness of the upper cut-off is evident in the latter figure. At 40 cps above the 88-kc carrier the discrimination has reached 20 db while the slope of the insertion loss versus frequency curve through this point is about 1 db per cps. Since at least two filters are connected in tandem in any program circuit a minimum of 40 db discrimination is provided to all frequencies in the unwanted sideband. The loss realized at frequencies outside the band also is shown in Fig. 2.

The delay distortion in the pass band of the filter, computed from the slope of its measured insertion phase characteristic, is given in Fig. 5. For short program systems, where no more than six filters are used in tandem, the delay distortion would not exceed the limits set for a high quality system. For longer systems it is necessary to equalize this delay distortion. The design and performance of the delay equalizers for this purpose are given in a separate paper.[3] These equalizers also include some attenuation equalization to correct for the systematic distortion of the filter.

Figure 6 shows an exterior view of the filter. On both sides of the mounting panel are metal containers which are provided with covers that can be soldered on to make a hermetic-sealed enclosure. In a corner of one can is a terminal box which contains the input and output terminals. These terminals are of the metal glass seal type which are vacuum tight. Mounted on brackets in each of the containers is a brass panel supporting the filter elements. One side of one of these panels is visible in Fig. 6, the other side is shown in Fig. 7.

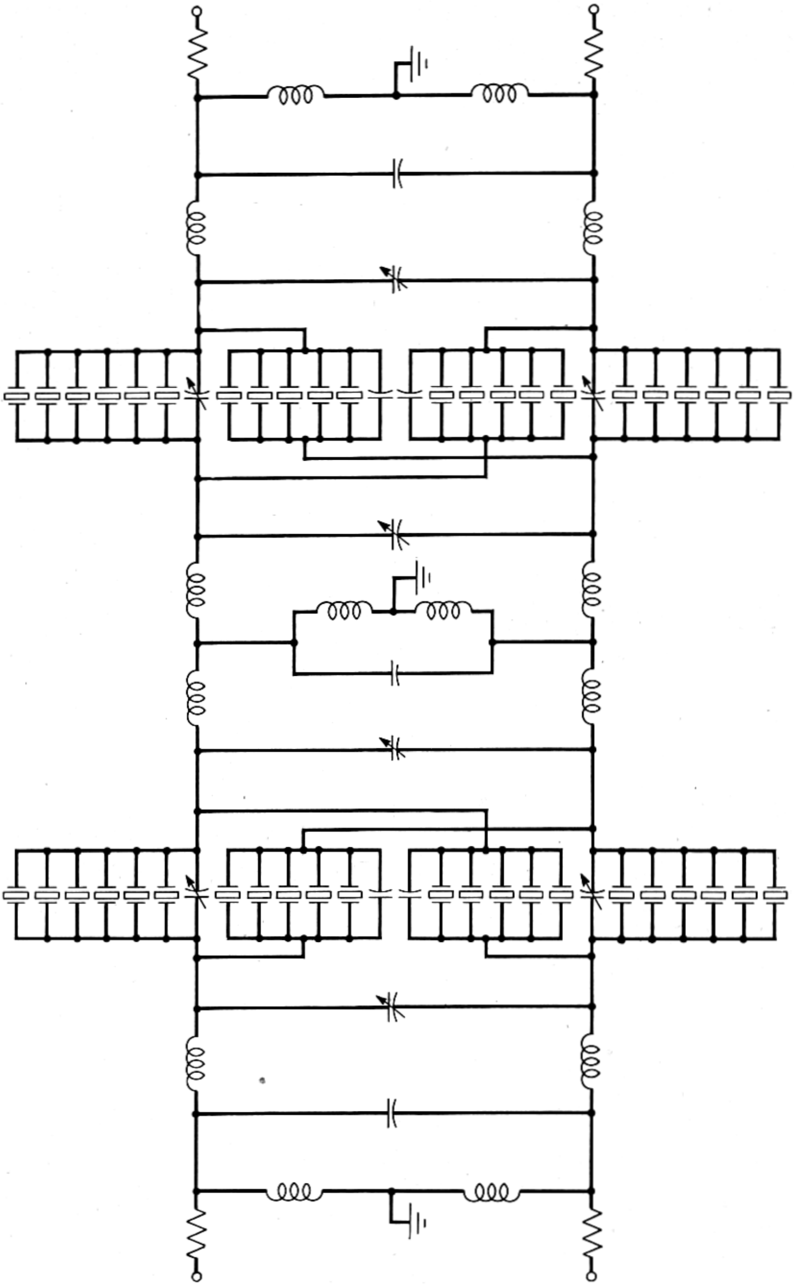

Fig. 1—Schematic of the channel selecting crystal band pass filter as constructed.

In Fig. 7 the two large cylindrical containers parallel to the panel house eleven of the balanced quartz crystal elements. The smaller cylindrical cans contain adjustable retardation coils while the rectangular cans house fixed coils. Adjustable air capacitors can be seen mounted on the hard rubber plate between the two crystal units.

The adjustment side of the brass panel is exposed in Fig. 6. Screwdriver adjustment of the retardation coils is possible through the circular holes at

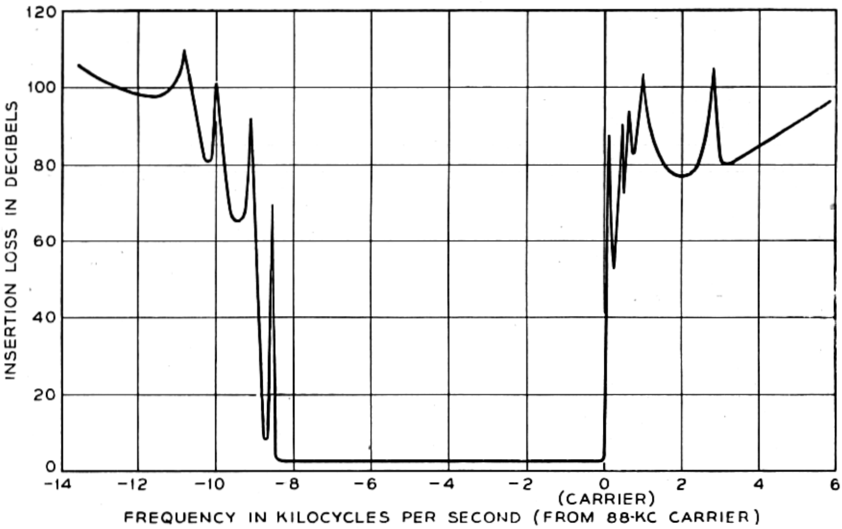

Fig. 2—The insertion loss-frequency characteristic of the filter.

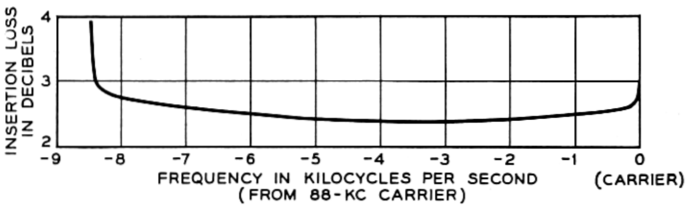

Fig. 3—Enlarged insertion loss-frequency characteristic of the filter pass band.

the top left and right of the panel. The rotors of three of the four air capacitors are visible inside the square cut-out in the panel. The panel in the lower half of the filter contains the remaining elements mounted and wired in a similar manner.

The schematic which was found to be most useful during the design of the filter is shown in Fig. 8. Thus the electrical circuit consists essentially of two complex lattice sections separated by one constant-k ladder section

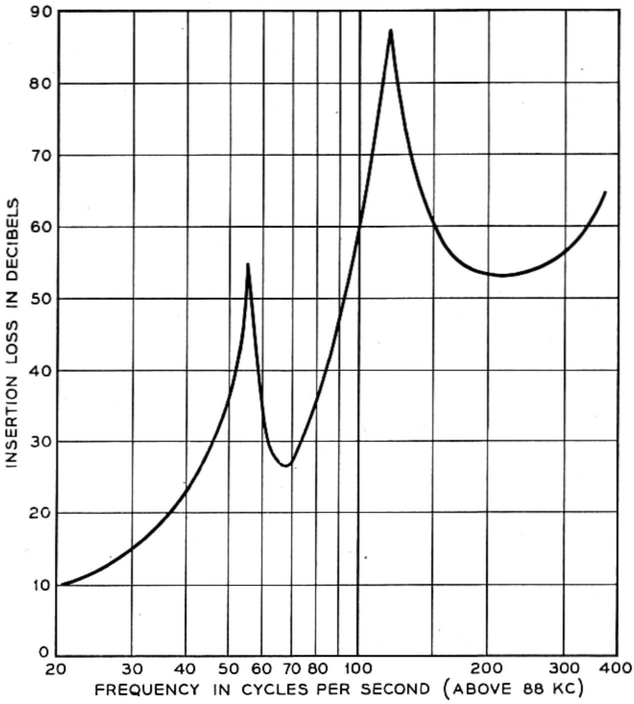

Fig. 4—The sharpness of the upper cut-off of the filter is shown in this enlarged loss characteristic.

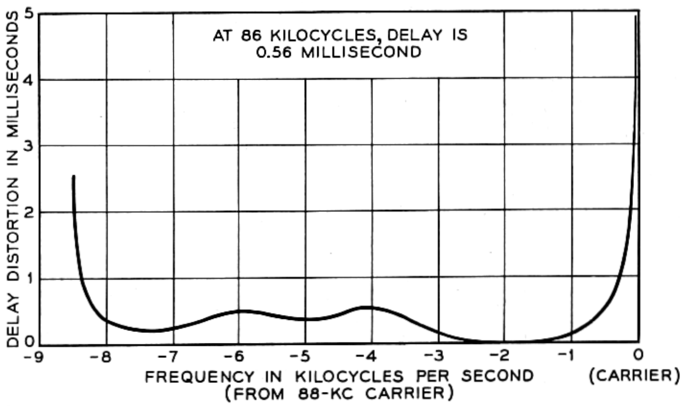

Fig. 5—Delay distortion in the pass band of the filter.

and terminated at each end by half-sections of the constant-k ladder type. The performance of the filter results almost entirely from the lattice sections since they control the flatness of the pass band, the sharpness of the cut-off and give practically all the discrimination required. It will be noted that the filter uses the equivalent of 130 electrical elements consisting of 63 inductors, 63 capacitors and 4 resistors.

The use of complex filter sections permits the realization of filter characteristics which have low distortion in the pass band and high discrimination outside the pass band with a more efficient utilization of elements than is possible with a larger number of simpler sections. Improved mathematical methods of network analysis developed in the past several years

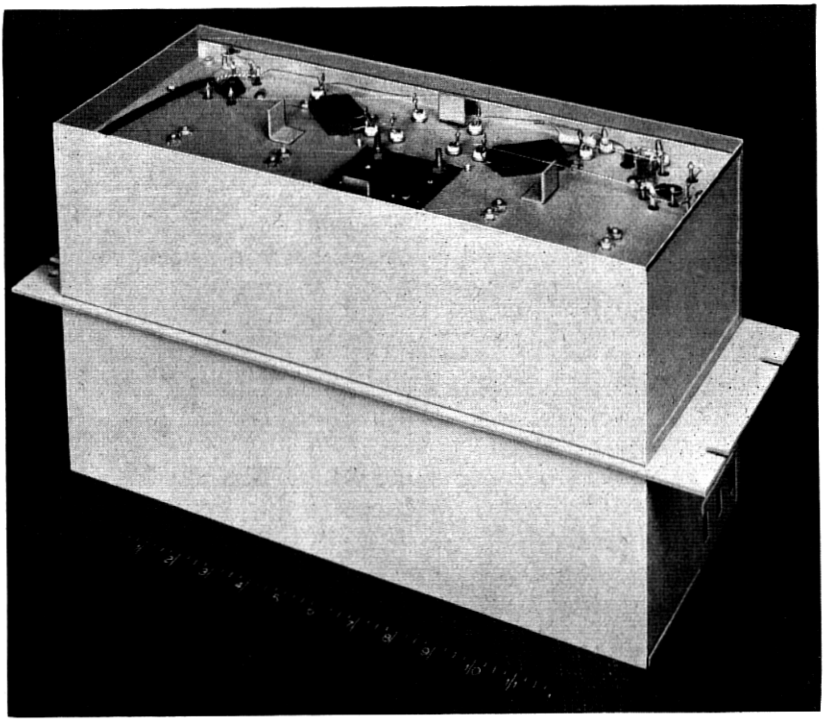

Fig. 6—Exterior view of the filter with one cover removed. After adjustments are completed the cover is soldered on to seal the assembly.

have made the design of such complex filter sections possible while recent developments of precise and stable filter elements and improved measuring circuits have made it possible to manufacture such filters.

It has been mentioned before that the use of quartz crystal elements restricts the filter band width which can be realized. In the frequency location selected for this filter (lower sideband of an 88-kc carrier frequency) a complex lattice section of the type shown in Fig. 8, when used alone, will permit the use of physical crystal elements for bands not over 7300 cps

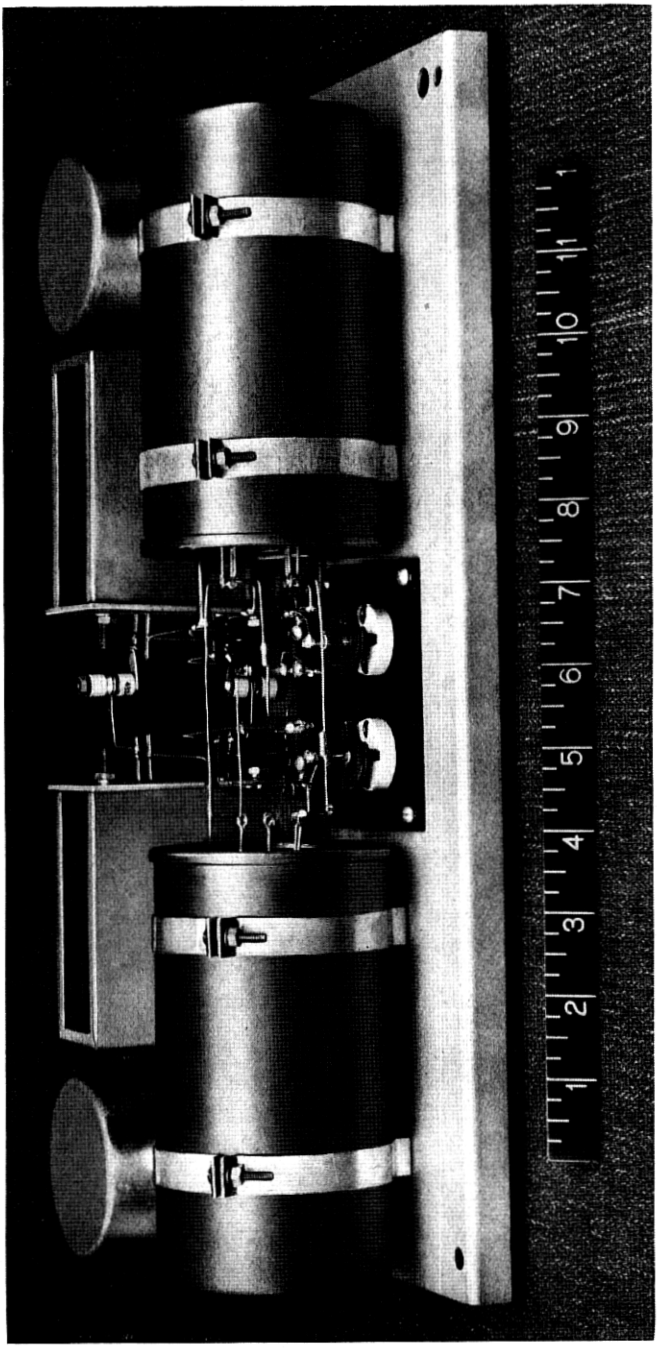

Fig. 7—Another view of one of the two panels which support the filter elements.

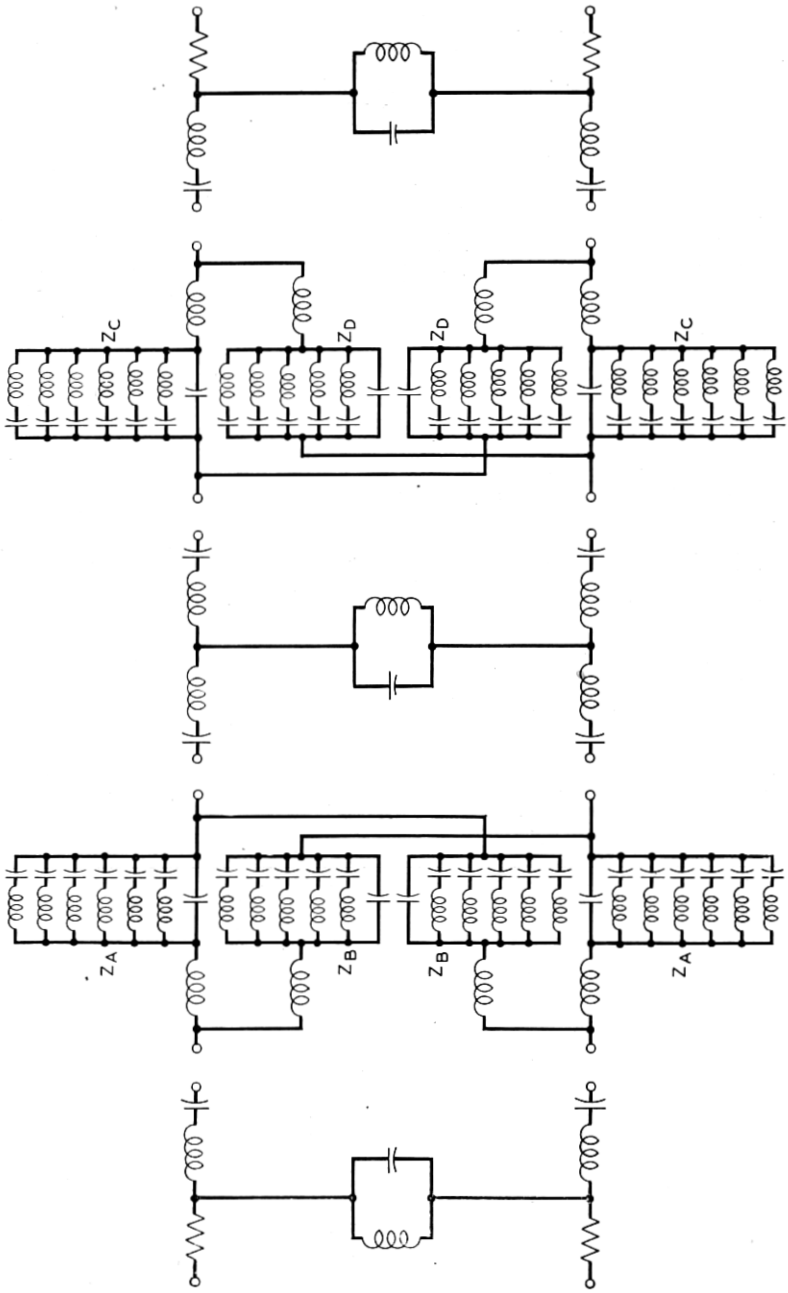

Fig. 8—The schematic used during the design of the filter contains 130 electrical elements.

wide. A wider band was obtained in this case by combining the complex lattice sections with ladder sections as shown in the figure. For this filter a combination of sections was designed which gave physically realizable crystal elements for a band width of 8.5 kc. This was the maximum band width possible without increasing the distortion in the band.

Summarizing, the filter design process consists of:

1. Design of wide band lattice sections which have quartz elements which cannot be realized in practice.
2. Design of electrical ladder sections of still wider band which introduce little distortion at the pass band frequencies of the lattice section. At this point the schematic is as shown in Fig. 8.
3. Combination of like elements, electrical transformations, and replacement of groups of elements consisting of an inductor and capacitor in series shunted by a second capacitor by their equivalent crystal elements. This gives the final schematic shown in Fig. 1, in which the crystal elements are physically realizable.

The general steps in the design of lattice filters[2,4] are as follows:

1. Choice of filter cut-offs.
2. Determination of number and location of impedance controlling frequencies to give a good match of image impedance to the termination.
3. Location of peaks of infinite attenuation to give the necessary transfer loss at frequencies removed from the pass band.
4. Determination of impedance level which gives the most reasonable element values.

Theoretically a filter could be designed which contains only one lattice section. The decision to split the filter into two sections was based on a desire to simplify the design to ease the manufacturing problems. The attenuation burdens of each section were reduced sufficiently to allow wider tolerances to be placed on the filter components. The last design steps are to determine the schematic of each section and to compute the theoretical element values in accordance with previously described methods.[2,4]

Although the filter elements computed were physically realizable they represented such extreme values as to introduce difficult problems. This was true especially of the crystal elements where the equivalent inductances of the eleven crystal elements in one section varied from 16 to 465 henries, a range of 1:29. A similar situation existed in the other section.

Crystal elements of the +5 degree X-cut type vibrating in their fundamental longitudinal mode are used in this filter. The equivalent inductance of such crystal elements varies directly with the thickness and inversely with the width of the plate. Therefore the high inductance plates are thick and narrow and the low inductance plates are thin and wide. In one section of the filter the dimensions of the plates required varied in width from

0.67 to 0.17 inch, in thickness from 0.119 to 0.012 inch and in length from 1.40 to 1.23 inches. The small variation in length is due to the fact that the length is determined primarily by the frequency of resonance of the plate and this change is small across the filter band. The temperature coefficient of the +5 degree X-cut quartz crystal element used in this filter is superior to the −18 degree X-cut longitudinal type which has been used in many other crystal filters but otherwise they are similar in use and in manufacture.

The filter attenuation distortion in the vicinity of the cut-offs is dependent on the dissipation in the elements which resonate there. In order to minimize this distortion, it has been found necessary to impose minimum Q requirements of 80,000 on the high-impedance crystal elements which resonate near the cut-offs. This high Q is realized by suspending the quartz crystal plates from fine wires[5] and operating them inside of evacuated containers. The low-impedance crystal elements which resonate at frequencies removed from the cut-offs require a minimum Q of 15,000. This comparatively low Q is realized by quartz crystal elements vibrating in air at atmospheric pressure.

In the equivalent electrical circuit of a quartz crystal element the large ratio of the shunt capacitance to the internal capacitance is a measure of the poor electromechanical coupling of quartz. For the +5 degree X-cut quartz crystal element this ratio of capacitances is about 140 for a plated blank before fabrication. It is obvious that fabrication, wiring and parasitic capacitances which may be in parallel with the quartz plate will make this ratio still higher and thus will reduce further the filter band width obtainable. For this reason it is important to keep to a minimum any capacitances which appear across any arms of the crystal lattices. One method used to minimize these capacitances was to assemble the eleven crystal elements required for each section in two containers instead of eleven separate ones. The five high-impedance elements requiring minimum Q's of 80,000 are assembled in one evacuated metal container and the six low-inductance elements having the lower Q's are assembled in another hermetic sealed container filled with dry air. A photograph showing the method of assembly used in given in Fig. 9.

A method was found to reduce the ratio of capacitances of the crystal elements. This method consists of dividing the plating on the surface of the quartz so that the driving voltage is removed from the end portions of the quartz plates. This plating division increases the equivalent inductance of the quartz plate but also decreases the direct capacitance between the plated surfaces. It has been found that the decrease in shunt capacitance with removal of plating is more rapid than the increase in equivalent inductance up to a certain point. If the plating is removed up to this optimum point it has been found possible to reduce the shunt capacitance about

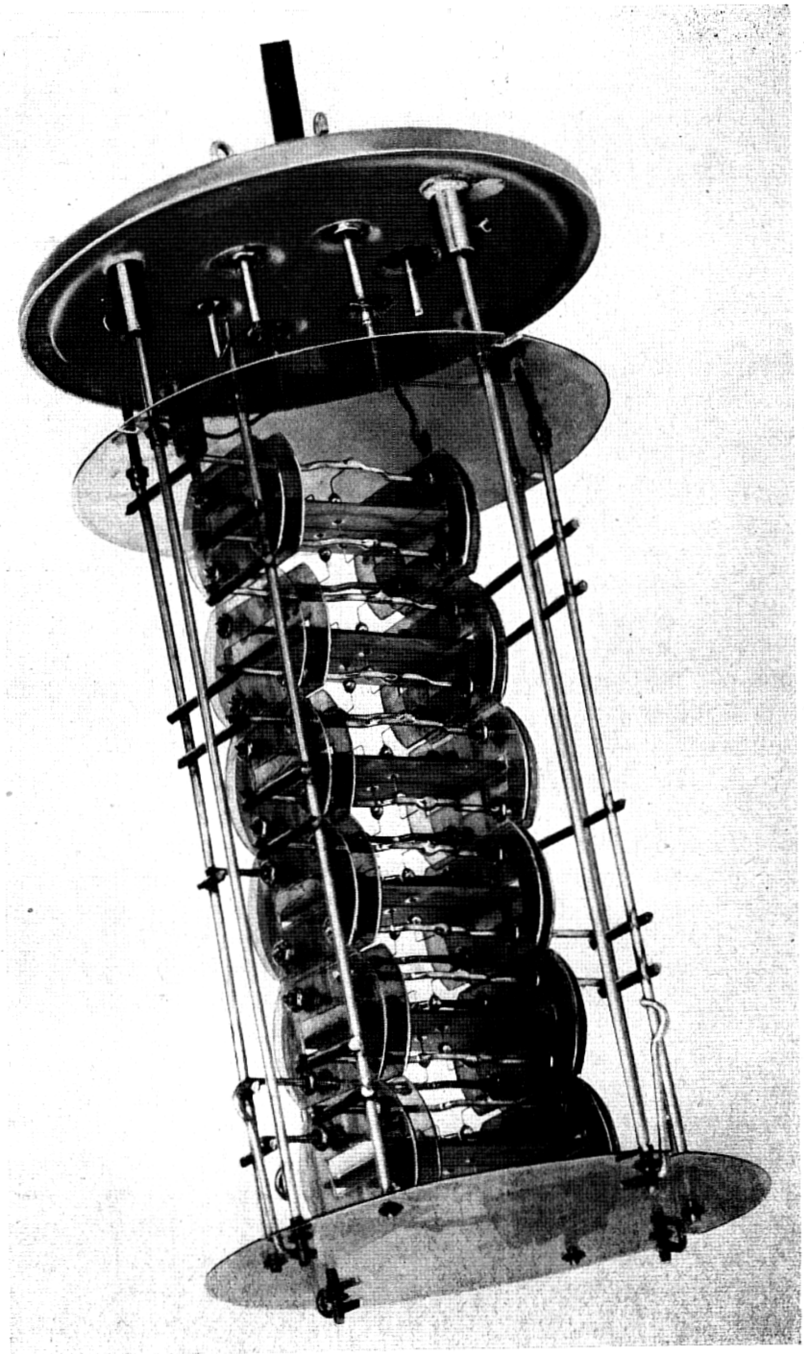

Fig. 9—Method of assembly of the quartz crystal elements.

17% below what it would be with a fully plated crystal element having the same inductance. This method of capacitance reduction was used on the six low-inductance crystal elements in each section. Another step in minimizing the unwanted capacitances was to design the retardation coils which connect to the terminal ends of each lattice to have as little capacitance as possible. Finally precautions were taken to keep the wiring capacitances to a minimum and the air condensers used inside the lattice for adjustment purposes are of special design having a minimum capacitance of only 0.5 mmf.

The resonant frequencies of each of the twenty-two crystal elements must be adjusted to the desired nominal frequencies within very close tolerances. On the ten high-impedance crystal elements the tolerance is ± 2 cps while on the 12 low-impedance crystal elements the tolerance is ± 5 cps. This precise frequency adjustment is accomplished by careful grinding of the length of the quartz plate.

The equivalent inductance of each of the 22 quartz crystal elements is required to be within two per cent of its nominal value. This specification is met primarily by close dimensional tolerances in the manufacture of the quartz plate. Any small adjustments which are necessary to meet this requirement are done by the aforementioned method of isolation of a small amount of plating from near the end of the quartz plate.

The four fixed retardation coils are adjusted to be within two per cent of their nominal inductance values. The variations from nominal are partially absorbed in the filter adjustment procedure where the coils are tuned with their associated variable capacitors to give the desired resonance frequency. The fixed mica capacitors are manufactured to be within 0.5 per cent of the desired nominal value. The three adjustable retardation coils are constructed to permit an inductance variation of five per cent on either side of their nominal values. This is done by moving a permalloy core in the field of the coil. Adjustment of these coils in the filter is accomplished by tuning them with their associated precision capacitor to give the desired resonance frequency within ± 25 cps. This type of adjustment procedure gives the correct LC product. The correct L/C quotient is obtained also since C is accurate to ± 0.5 per cent. The two resistors at each end of the filter compensate for the dissipation in the end retardation coils and thus restore the terminating impedance to the value required for optimum filter performance.

Each lattice of crystal elements and capacitors is a four-terminal bridge which is adjusted for maximum bridge balance at a particular frequency by means of the variable air capacitors in two of the arms. The precision of inductance adjustment of the crystal elements insures that the other peaks of attenuation will be sufficiently close to their nominal locations.

To obtain maximum loss at the filter peaks it is necessary to secure a conductance balance in each lattice section as well as a susceptance balance. This can be done if care is exercised in the choice of materials used in fabricating the crystal elements and capacitors which appear inside the lattice. In this case the crystal element insulators and dielectrics consist of glass, mica, quartz and clean dry air or vacuum while the air capacitors use glass, ceramic and air for their insulators and dielectrics. If these materials are clean and dry they have very low conductance and do not influence the bridge balance. A complete discussion of the effects of impedance unbalances on crystal lattice performance has been given in a recent paper by E. S. Willis.[6]

To further insure that dirt and moisture will not influence its performance the filter is adjusted, tested and hermetically sealed in an air conditioned room where the relative humidity does not exceed 40 per cent. Since manufacture started about the beginning of 1946 several hundred of these filters have been made and are functioning satisfactorily in the telephone plant.

Band Elimination Filter at Branching Points

When broad-band carrier systems are equipped for the transmission of a carrier program channel, it is frequently necessary to provide between carrier terminals intermediate or branching points at which the program may also be received. If only receiving facilities are involved, rather simple bridging arrangements can be provided. However, program network needs often require a more flexible arrangement at the branching point so that a line may be cleared of the program originating at one terminal and a new program introduced for transmittal toward the next terminal.

To do this without affecting the message channels also being transmitted on the line, a filter has been developed to block the program channel already on the line while freely transmitting the message channels. With this filter in the circuit the high-frequency line between the branch point and the following terminal is free of program frequencies and the program originating at the branch point may be sent toward that terminal.

Since the program channel occupies frequency space near the center of the 12-channel message group, the remaining message channels appear above and below the program frequencies. Therefore the blocking filter at the branching points must be of the band elimination type. The circuit employing this filter may be designed to block either at line frequencies or at basic group frequencies. The latter method, of course, requires that a demodulation process be provided to translate line frequencies to basic group frequencies before the blocking filter is inserted in the circuit.

The band elimination filter described herein was developed for the type K carrier system (Carrier-on-Cable) for which the first option mentioned above was chosen. This filter operating at the line frequencies of the type K system is required to transmit frequencies from 12 to 31.6 kc and 44.2 to 60 kc while blocking those from 32 to 43.2 kc. Actually the filter will transmit frequencies below 12 kc and above 60 kc but these do not appear on the type K line and therefore there are no requirements in these ranges.

The filter which performs these functions is shown schematically in Fig. 10. Several factors made its design difficult. A high level of discrimination of the order of 75 db is required over a wide frequency range of about 12 kc. Also the allowable waste interval between wanted and unwanted frequencies is very small. The filter must transmit with a maximum distortion of 0.2 db to within 97.5% of the first unwanted frequencies at which a discrimination level of 75 db is required.

Because of the severe requirements the familiar image parameter design method was not employed. In this, as is well known, the composite filter first presented by Zobel[7] is made up of sections with matched image impedances but different transfer constants depending upon the attenuation requirements. Instead, it was felt that a design method proposed by Darlington[8] offered a better possibility of meeting the requirements with a reasonably sized filter. This procedure known as the *insertion loss* method is based upon the determination of a four-terminal transducer of reactances which, when inserted between definite resistance terminations, will produce a specified loss characteristic. A filter so designed has an advantage over image parameter filters in that the attenuation obtainable is greater for the same effective cut-off and an equal number of elements. *Effective cut-off* as used here means the last frequency of interest in the transmitted band. It is possible, therefore, with an *insertion loss* filter to use fewer elements for a given attenuation, or to obtain a wider transmission band with the same number of elements.

The advantage inherent in the newer design method is not derived from a difference in structure. In configuration there is no way to distinguish such a filter from one of conventional image design. The difference lies solely in the element values. A simple way to visualize how the *insertion loss* design varies from image design is to consider that the newer method removes an arbitrary restriction placed upon the image theory to simplify the mechanics of design. The restriction is that the nondissipative image attenuation must be identically zero over continuous frequency ranges including the transmitted bands and other than zero everywhere else. This leads to the familiar ladder image filter composed of matched sections, or the lattice filter with coincident critical frequencies.

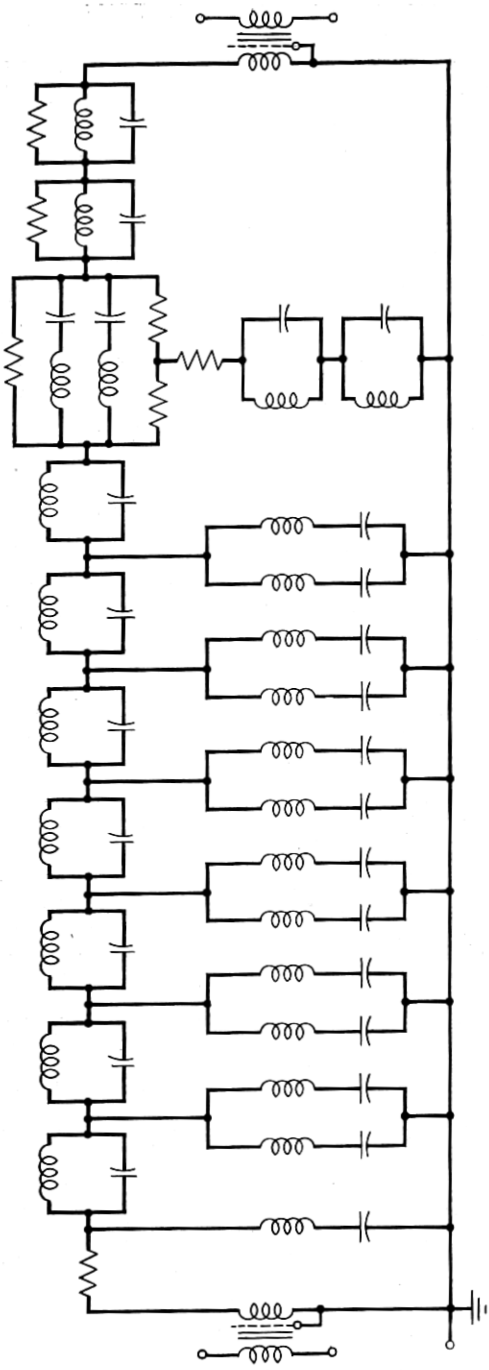

Fig. 10—Schematic of the band elimination filter used at branching points of the type K system.

Analysis of an *insertion loss* ladder filter shows that it may be considered a composite of image sections which are not matched in image impedance. As a composite filter this means that the effective pass band has been split into a number of pass bands each separated by a small attenuation region. Darlington has formulated the process by which these bands can be so arranged that advantage can be taken of the fact that the image attenuations in these bands for small mismatch are comparable to the terminal effects and that reflection gains up to 6 db are possible in the same regions. The combination of these effects, which can be controlled up to and including the cut-off, gives the *insertion loss* filter its improved performance since the *effective* and *theoretical* cut-offs can be made identical, with no

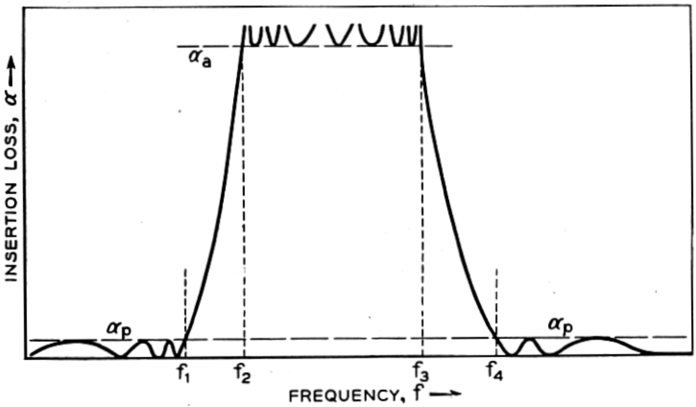

Fig. 11—Non-dissipative filter characteristic obtained by use of Tchebycheff parameters in pass bands and attenuation band.

frequency space needed for the rounding due to the terminal effects in image filters.

In general the mathematical steps required to design a filter by this method are as follows: An insertion loss frequency function is chosen which will satisfy the filter requirements and will lead to a structure economical of elements. From this are found the open and short circuit impedances of the proposed network which is normally of the standard lattice or ladder forms. Finally from these expressions the element values are determined.

The particular form of *insertion loss* design employed for the filter described here is a special case of the general theory. The filter requirements lent themselves to the use of Tchebycheff parameters simultaneously in the pass bands and attenuation band. The application of these parameters was first described by Cauer.[9] The typical non-dissipative characteristic resulting from their use is shown on Fig. 11. It is seen that the pass band characteristic is of the *ripple* type with equal maxima and equal minima. In the attenuation region the valleys of loss are of equal value.

The general form for the insertion power ratio to obtain the desired characteristic is:

$$e^{2\alpha} = \frac{4R_1 R_2}{(R_1 + R_2)^2} [1 + (e^{2\alpha_p} - 1) \cosh^2 \theta_I]$$

In this equation R_1 and R_2 are the resistive terminations and α_p is the maximum ripple in the pass band as shown in Fig. 11.

θ_I represents a function of frequency so chosen that $\cosh \theta_I$ is an odd or even rational function of frequency. Also θ_I must be a pure imaginary throughout the passing band and must be of the form $(\alpha_I + n\Pi i)$ in the attenuation region. The term α_I is real at all attenuation frequencies becoming infinite at those required by the specification of minimum α_a in Fig. 11.

Darlington further showed that θ_I closely conforms with the image transfer constant of an image parameter filter if the effective pass band of the insertion loss filter coincides with the theoretical pass band of the image filter. Based on this conclusion a design method was formulated which permits a reference filter derived from image parameters to be used as the basis of the *insertion loss* filter. There is, of course, no correspondence between the elements of the reference image filter and the insertion filter. This reference filter is not a requisite to the development of the insertion theory but it does offer a convenient and well known transfer constant which is the right functional form for use in the insertion power ratio stated above.

Referring again to Fig. 11, the approximate minimum loss, α_a, determines the number of peak sections required in the reference filter from the relationship:

$$\alpha_a = 20 \log (e^{2\alpha_p} - 1) - 10(2m + 1) \log q - 18$$

where "m" is the number of peaks required and α_p is the band ripple function as before. The new term introduced here is "q" which is directly tied up with the selectivity demanded of the filter, i.e., the amount of frequency space avilable between the last useful frequency or *effective* cut-off and the first frequency at which attenuation equal to α_a is needed. The relationships are as follows:

$$q = \frac{1}{2}\left[\frac{1 - \sqrt{K'}}{1 + \sqrt{K'}}\right] + \frac{1}{16}\left[\frac{1 - \sqrt{K'}}{1 + \sqrt{K'}}\right]^5$$

where $K' = \sqrt{1 - K^2}$

and $K = \dfrac{f_3 - f_2}{f_4 - f_1}$

The filter described here actually consists of two filters connected in tandem, each derived from a different power ratio. This step was taken because of the relatively low dissipation factor realizable with coils of reasonable size. By dividing the total attenuation between two power ratios, lower overall distortion due to dissipation was achieved. The distortion represented by the non-dissipative ripple "α_p" was minimized by so assigning the frequencies of infinite attenuation to the two functions that phasing in of the ripples was avoided as far as possible.

The two power ratios selected are:

$$e^{2\alpha_1} = 1 + (e^{2\alpha_p} - 1)\cosh^2 \theta_{I_1}$$

$$e^{2\alpha_2} = \frac{4R_1 R_2}{(R_1 + R_2)^2} [1 + (e^{2\alpha_p} - 1)\cosh^2 \theta_{I_2}]$$

For these the peak frequencies were assigned on an alternate basis as follows:

$$\text{To } \theta_{I_1} : m_1 , m_3 , m_5 \text{ and } m_7$$

$$\text{To } \theta_{I_2} : m_1 , m_2 , m_4 \text{ and } m_6$$

with the value of "m" decreasing from m_1 to m_7 and $m_1 = 1$. The parameter "m" has the same meaning as in image filter theory.

The next step in the process is the finding of the roots of the two power ratios. These may be obtained from the following expansions:

For $e^{2\alpha_1}$ representing a reference filter of $3\frac{1}{2}$ sections:

$$(m_3 + x)^2(m_5 + x)^2(m_7 + x)^2(1 + x) + \left(\frac{e^{\alpha_p} - 1}{e^{\alpha_p} + 1}\right)(m_3 - x)^2(m_5 - x)^2(m_7 - x)^2(1 - x) = 0$$

which is expressed in the form

$$K_1 [x^7 + a_1 x^6 + a_2 x^5 + a_3 x^4 + a_4 x^3 + a_5 x^2 + a_6 x + a_7 = 0]$$

For $e^{2\alpha_2}$ representing a reference filter of 4 sections:

$$(m_2 + x)(m_4 + x)(m_6 + x)(1 + x) + i\sqrt{\frac{e^{\alpha_p} - 1}{e^{\alpha_p} + 1}}(m_2 - x)(m_4 - x)(m_6 - x)(1 - x) = 0$$

which is expressed by

$$K_2 [x^4 + a_8 x^3 + a_9 x^2 + a_{10} x + a_{11} = 0]$$

In the above expressions $x = \sqrt{1 + \frac{1}{p^2}}$ where $p = i\omega$ and α_p for the filter discussed here is 0.1 db. From the roots obtained from the above equations of 7th degree and 4th degree complexity, the open and short-circuit impedances are determined which in turn lead to the element values. The complete development of the process resulted in the filter portion of the network shown on Fig. 10.

The remainder of the schematic shows the equalizer which corrects the rounding of the filter characterisic near the cut-offs due to dissipation. The equalizer is of conventional bridged "*T*" design with constant "*R*" impedance in tandem with a simple series section.

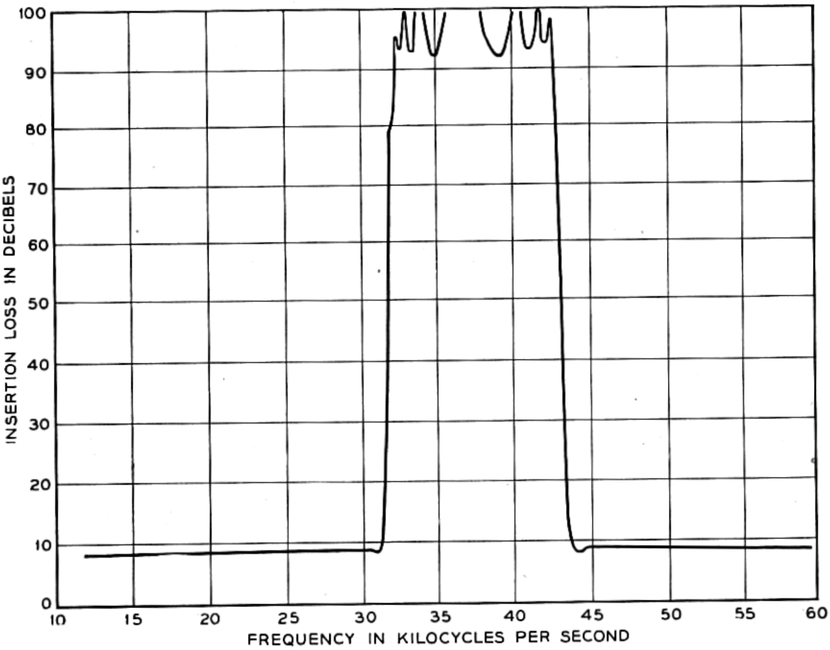

Fig. 12—Insertion loss-frequency characteristic of the band elimination filter.

Repeating coils are required as shown because the filter was designed at a 600-ohm level to give commercial elements whereas it is required to operate between 135 ohm resistances. In the schematic a resistance will be noted in series with one termination. This is needed because the "insertion" design with inverse impedance terminations as shown here requires unequal terminations to produce the specified loss characteristic. Usually this would be taken care of by proper design of the repeating coil but, in this case, economic reasons dictated the use of the same repeating coil at both ends of the structure. The termination was therefore built out with a physical resistance. This of course introduces a flat loss but in this case enough gain was available in the circuit to permit it.

On Fig. 12 is shown a typical transmission characteristic when the filter is operating between 135-ohm resistances. A variety of component parts are required to give this performance. The filter portion employs mica condensers throughout and a mixture of molybdenum permalloy and air-core retard coils. As many permalloy coils are used as possible in order

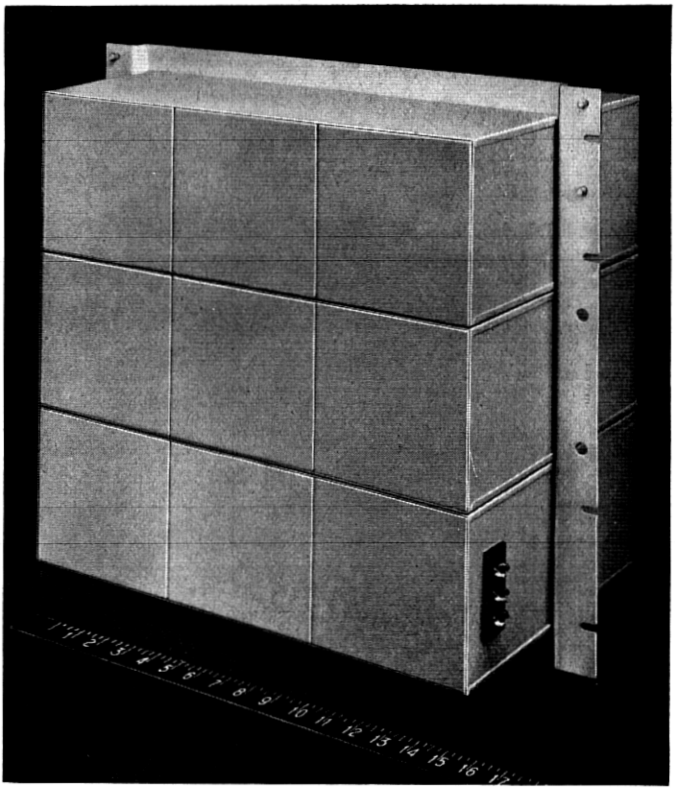

Fig. 13—Exterior view of the band elimination filter.

to obtain high "Q". The air-core coils, of an adjustable type, are used in those arms which control the peak frequencies near the pass band. These arms must be adjusted very accurately for resonance in order to maintain the steep slope of loss in the cut-off region. The equalizer sections employ duolaterally wound air core coils also adjustable in order to set the pass band losses accurately. Mica and paper condensers are used in the equalizer, the latter being used where capacity values make mica condensers extremely expensive.

A completed filter is shown in the photograph of Fig. 13, while the internal arrangements of one portion of the assembly are shown on Fig. 14.

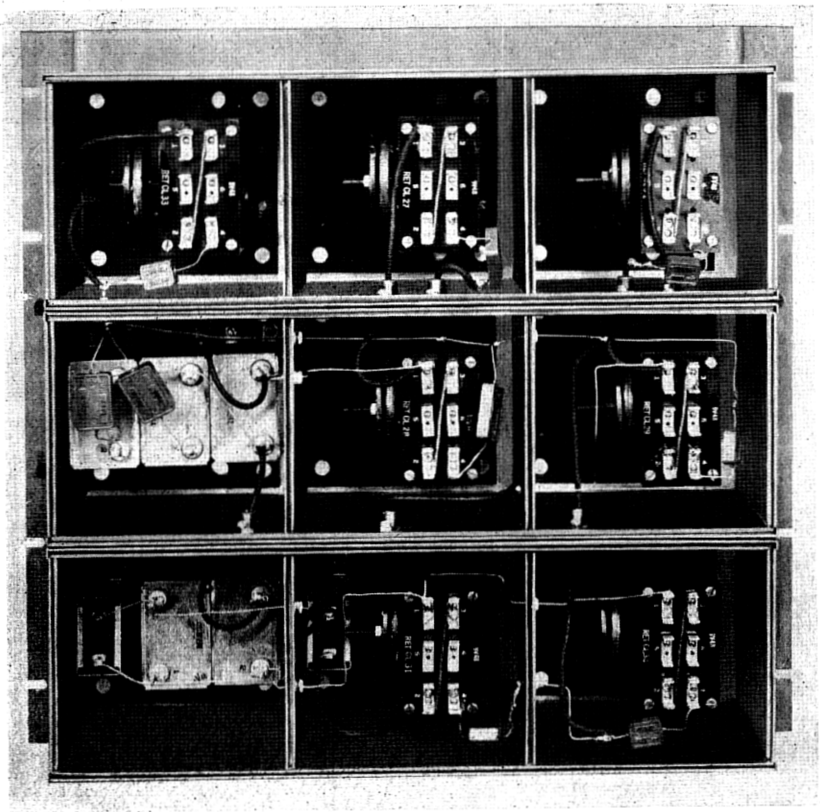

Fig. 14—Interior view of one portion of the assembly of the band elimination filter.

Phase Simulating Network

When program rearrangements at a branching point are required, the band elimination filter must be switched into or out of the through transmission path. This transfer is accomplished without opening the through path. Thus, for a brief time during the switching interval, message channels are transmitted simultaneously through the filter and the non-blocked circuit. A large phase difference between the two parallel paths is introduced by the filter which, in the absence of phase correction in the through circuit, could cause errors in the transmission of voice frequency telegraph signals. Therefore a network having phase shift similar to that of the filter over most of the message range is provided in the through circuit.

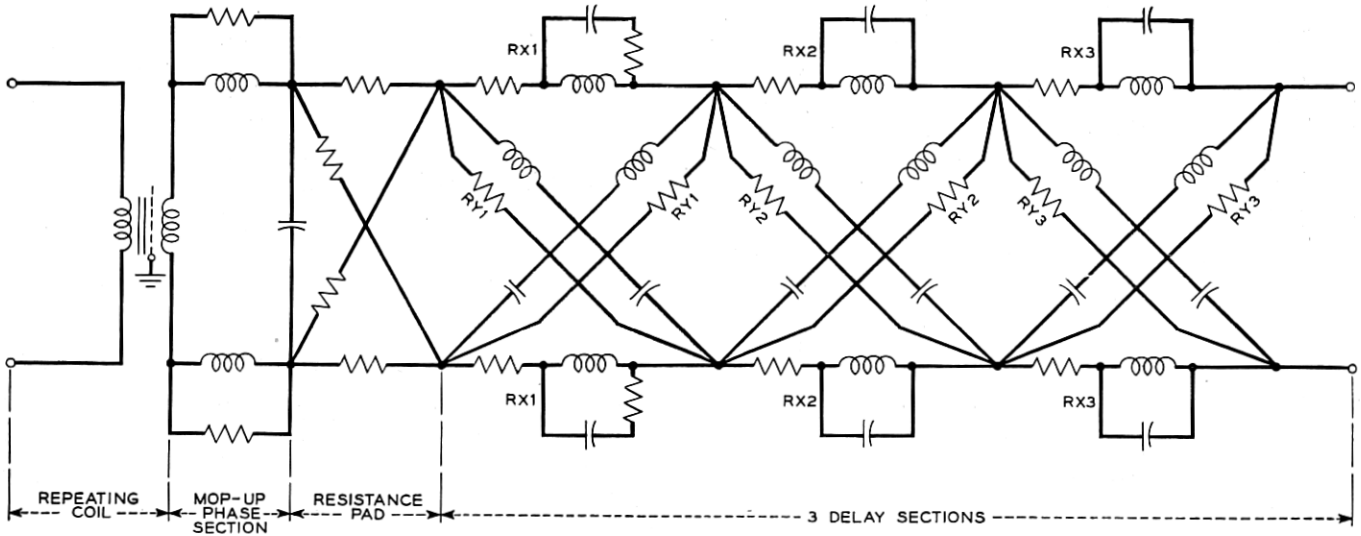

Fig. 15—Schematic of the phase simulating network showing the component sections.

The phase simulating network is shown in schematic form in Fig. 15. The network is a balanced structure and consists of the following pieces of apparatus connected in tandem:

1. An input repeating coil to improve the longitudinal balance at the sending end,
2. A half-section high-frequency cut-off low-pass filter to mop up the phase shift introduced by two repeating coils of the band elimination filter,

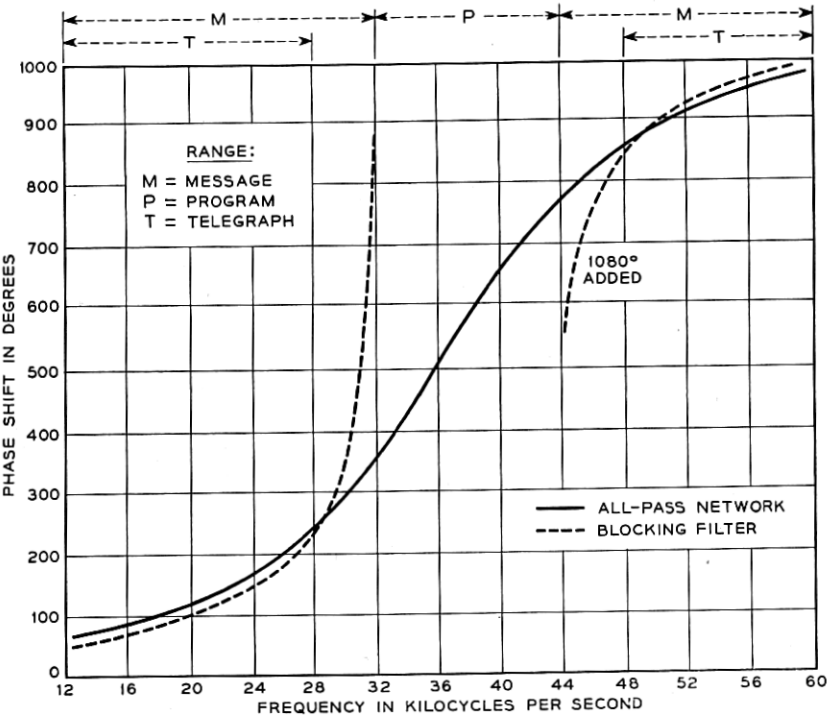

Fig. 16—Phase shift-frequency characteristic of the phase simulating network.

3. A resistance pad to equalize the over-all loss level of the all pass network to within ±0.1 db of the pass band loss of the band elimination filter, and
4. Three delay sections, self equalized for loss,[10] for simulating the phase shift of the band elimination filter.

The network simulates the phase shift of the band elimination filter over the frequency ranges covered by message channels 1 to 4 and 10 to 12 to within 20 electrical degrees as shown in Fig. 16. As phase simulation is incomplete in the frequency ranges occupied by channels 5 and 9 due to the steep phase shift slope of the band elimination filter near its cut-off points, no telegraph channel are assigned to these channels of type "*K*" carrier circuits equipped with branching points.

The phase shift of the band elimination filter is discontinuous between its cut-off frequencies and has a positive slope with frequency in its pass bands. As the phase shift of a delay section increases continuously with frequency, it is impossible to provide the exact counterpart of the filter in a delay network. However, the addition of any multiple of 2Π radians does not change the transmission characteristic. Hence 6Π radians (3

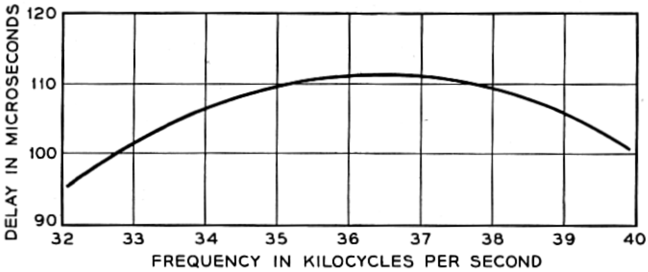

Fig. 17—Delay of the phase simulating network at program frequencies.

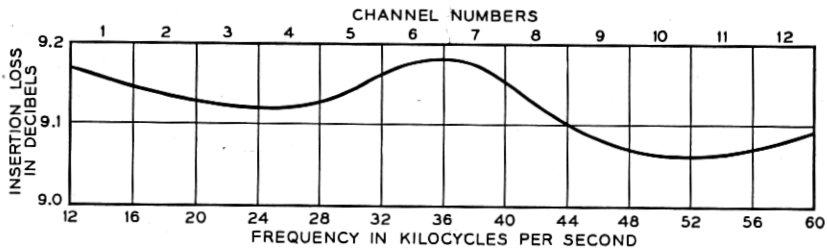

Fig. 18—Insertion loss-frequency characteristic of the phase simulating network.

revolutions) are added to the phase shift of the elimination filter above the upper cut-off to simulate its phase characteristic in the 10 to 12 message channel range, as well as to provide an almost linear phase slope in the 32 to 40 kc program channel range resulting in minimum delay distortion.

The delay distortion of the network over the program channel is approximately 16 microseconds as shown in Fig. 17. The loss distortion over the program channel is approximately 0.05 db and over any one message channel it is less than 0.05 db as shown in Fig. 18.

The self loss equalizing feature of a delay section is evaluated at zero frequency in the form of a resistance pad by making the pad loss approximate the insertion loss at the critical frequency of the delay section. The resistance R_x located in the series branch may be evaluated from the expression

$$R_x = R\frac{\epsilon^\theta - 1}{\epsilon^\theta + 1} - R_{DC}$$

in which ϵ^θ is the transfer loss in nepers at the critical frequency, R is the 135 ohm resistance termination and R_{DC} is the DC resistance of the inductance coil. The resistance Ry located in parallel with the series resonant branch may be evaluated from the expression $Ry = \frac{R^2}{R_x}$.

By changing the loss, ϵ^θ, of the derived resistance pad at zero frequency slightly from the measured loss at the critical frequency of the delay section, a suitable loss compensation may be realized to produce an optimum loss equalization over the message and program channel ranges. It is satisfactory to follow this technique when the condenser Q factor is much greater than the coil Q factor. When this condition exists, the insertion loss about the critical frequency becomes geometrically dissymmetrical, that is, the loss falls off more rapidly for frequencies above the critical frequency because of the controlling condenser Q factor.

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