EXTRACTS FROM COURSE BOOKS

FOR SECOND YEAR TRAINEES.

PART 1. PADS, FILTERS AND EQUALISERS.
(L.L.E. 1, Paper 3, Pages 1-16).

PART 2. V.F. REPEATERS & HYBRID COIL.
(L.L.E. 2, Paper 1, Pages 1-8).

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COURSE OF TECHNICAL INSTRUCTION.

NETWORKS, ATTENUATORS, FILTERS AND EQUALISERS.

CONTENTS:

1. INTRODUCTION.

1.1 There is an almost infinite number of ways of connecting resistances, inductive reactances and capacitive reactances together to perform different functions, the different circuits so formed being called Networks. From the multiplicity of different possible combinations, a few types are encountered again and again in communication circuits so frequently as to warrant special treatment. It is the purpose of this Paper to study two of the most frequently encountered types and to deal with some of their applications.

1.2 The two types to be studied are called T and π networks from their configuration. These networks may be balanced or unbalanced, as shown in Fig. 1.

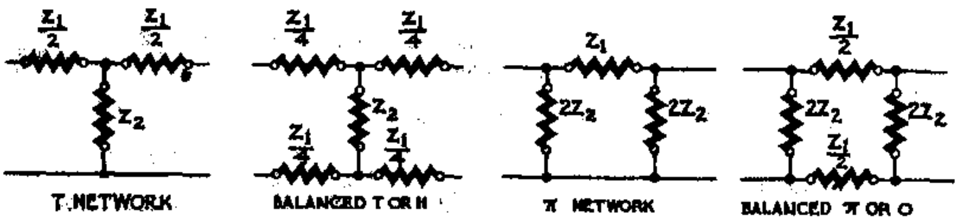


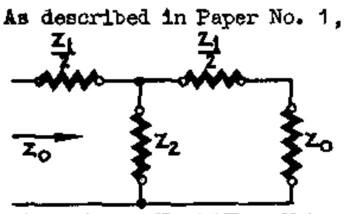
FIG. 1. T AND π SECTIONS.

2. CHARACTERISTIC IMPEDANCE OF NETWORKS.

2.1 If an infinite number of the single networks, or sections, shown in Fig. 1 were connected together and the input impedance measured or calculated, as was done in Paper No. 1, the same result would be obtained. that is. the infinite number of sections would exhibit a finite input impedance. This impedance is frequently referred to as the Iterative Impedance (to iterate means to repeat). In these books, this impedance is called the Characteristic Impedance and designated Z_o in order to be consistent with Paper No. 1 on Infinite Lines.

3. CHARACTERISTICS OF T SECTIONS.

3.1 As described in Paper No. 1, terminating a T network or section in its characteristic impedance makes the input impedance of the network equal to the characteristic impedance. This is shown in Fig. 2.



T SECTION TERMINATED IN Z_o.

FIG. 2.

3.2 Characteristic Impedance of T Sections. The characteristic impedance of such a network is:-

$$Z_o = Z_1Z_2 + \frac{Z_1^2}{4} \quad \text{.........................(1)}$$

3.3 Attenuation Produced by T Sections. The attenuation of a T network can be calculated from the ratio of the input and output currents. These currents are shown in Fig. 3.

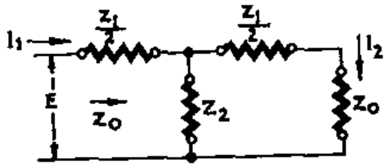


FIG. 3. CURRENTS IN T SECTION.

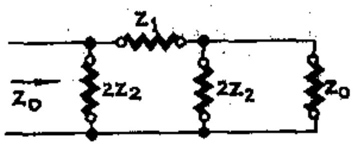
The ratio of input and output will be $\frac{I_1}{I_2}$, and is given by -

$$\frac{I_1}{I_2} = \frac{Z_2 + \frac{Z_1}{2} + Z_o}{Z_2} \quad \text{.......................(2)}$$

$$\text{or } \frac{I_1}{I_2} = 1 + \frac{Z_1}{2Z_2} + \frac{Z_o}{Z_2} \quad \text{......................(3)}$$

4. CHARACTERISTICS OF π SECTIONS.

4.1 Characteristic Impedance of π Sections. In Fig. 4, Z_o is in parallel with $2Z_2$, this combination being in series with Z_1, and this combination, in turn, being in parallel with $2Z_2$. The characteristic impedance of such a network is given by -



π SECTION TERMINATED IN Z_o.

FIG. 4.

$$Z_o = \frac{Z_1Z_2}{\sqrt{Z_1Z_2 + \frac{Z_1^2}{4}}} \quad \text{.........................(4)}$$

$$\text{or } Z_o = \frac{Z_1Z_2}{Z_{OT}} \quad \text{................................(5)}$$

where Z_{OT} is the characteristic impedance of a T section, having Z_1 as its series impedance and Z_2 as its shunt impedance.

4.2 Attenuation of π Network. This is given by -

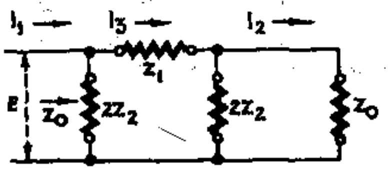


FIG. 5. CURRENTS IN π SECTION.

$$\frac{I_1}{I_2} = 1 + \frac{Z_o}{Z_2} + \frac{Z_1}{2Z_2} + \frac{Z_1Z_o}{4Z_2^2} \quad \text{.................(6)}$$

/4.3

4.3 T Sections are frequently called "mid-series" sections, and the impedance of such a section is referred to as the "mid-series impedance." π sections, on the other hand, are frequently called "mid-shunt" sections, and the impedance of such a section is referred to as the "mid-shunt impedance." The reason for these designations is shown in Fig. 6, which is termed a "ladder" network because of its configuration.

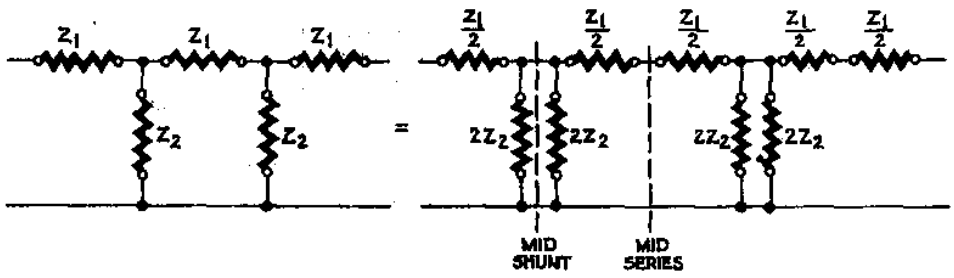


FIG. 6. MID-SERIES AND MID-SHUNT DERIVATION.

5. PADS AND ATTENUATORS.

5.1 Sometimes it is necessary to reduce the amount of power input to, or output from, a piece of equipment by a fixed amount. This is done by means of a suitable network having the desired fixed loss, such a network being termed a Pad.

5.2 Again, the same item of equipment, for example, an oscillator, may be used under circumstances which require the output power to be varied by known and variable amounts. This is done by means of a suitable network having the desired variable loss, such a network being termed an Attenuator.

5.3 Pads and Attenuators are usually of the "T" or "Balanced T" type, depending on whether strict balance is necessary or not. The series and shunt elements are purely resistive, as they usually have to offer a loss or attenuation which is the same at all frequencies. Further, the pads and attenuators usually operate between similar impedances, that is, their output and input impedances are the same, such networks being termed Symmetrical Networks. Where the input and output impedances of a network are unequal, the networks are termed Unsymmetrical Networks.

5.4 The problem of designing a symmetrical pad thus becomes one of working out values for the resistances in Figs. 2 and 4, so that -

$$\sqrt{R_1R_2 + \frac{R_1^2}{4}}$$

produces the required impedance between which the pad must work, together with the required amount of attenuation. In the above expression, R_1 and R_2 replace Z_1 and Z_2, respectively, of equation (1) because, as mentioned above, the pad elements are purely resistive. Tables are generally available where networks have to be designed frequently, such tables simplifying the calculations. Table 1 is included to indicate the accuracy required. By applying equations (1) or (4), it will be found that the characteristic impedances very closely approximate 600 ohms, whilst an application of equations (3) or (6) will show that the attenuation is correct.

/Table 1.

	T SECTION		π SECTION	
Loss in db.	Total Series Resist.	Shunt Resist.	Total Series Resist.	Shunt Resist.
1	69 ohms	5,208 ohms	68.6 ohms	10,440 ohms
2	137.6 ohms	2,582 ohms	139.4 ohms	5,232 ohms
3	205.4 ohms	1,703 ohms	212.5 ohms	3,505 ohms
4	271.6 ohms	1,249 ohms	287.5 ohms	2,651 ohms
5	336.2 ohms	987.6 ohms	364.5 ohms	2,141 ohms

VALUES OF RESISTANCES FOR 600 OHM ATTENUATORS.

TABLE 1.

5.5 The problem of attenuators is frequently complicated by the fact that the input and output impedances of the attenuator must remain unchanged as the values of the resistances in the attenuator change to produce the required amount of attenuation.

6. CHARACTERISTIC IMPEDANCE OF T NETWORKS OF PURE REACTANCES.

6.1 In order to understand the operation of filters, a knowledge of how networks having purely reactive elements behave is necessary.

6.2 As examples, some T networks of reactances will be examined. Figs. 7a and 7b show T sections of pure inductance and capacitance, respectively. The characteristic impedance of a T network of pure inductances is an inductive reactance at all frequencies, and the characteristic impedance of a T network of pure capacitances is a capacitive reactance at all frequencies.

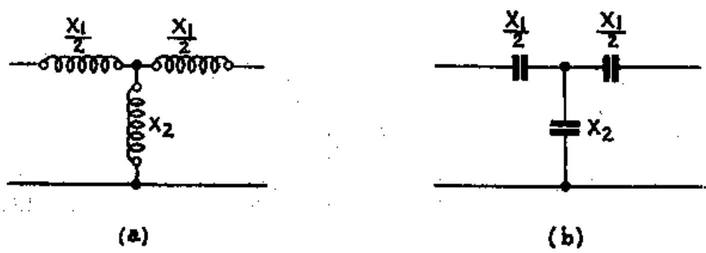


FIG. 7. T NETWORKS WITH SIMILAR REACTIVE ARMS.

6.3 Figs. 8a and 8b show two other arrangements of T sections, these having opposite types of reactances in the arms. The type of network shown in Fig. 8a has a characteristic impedance which is purely resistive at frequencies from zero up to that at which $X_2 = \frac{X_1}{4}$, and above that frequency the characteristic impedance is a pure inductive reactance. The network of Fig. 8b has a characteristic impedance which is a pure capacitive reactance between zero frequency and that at which $X_2 = \frac{X_1}{4}$. Above this frequency, the characteristic impedance is a pure resistance.

/Fig. 8.

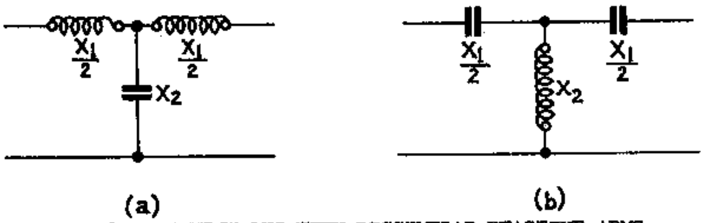


FIG. 8. T NETWORKS WITH DISSIMILAR REACTIVE ARMS.

6.4 It is not possible to explain by physical means, as was done for Wave Propagation in Paper No. 1 of this book, how a network consisting of pure reactances can behave as a pure resistance over a certain range of frequencies and as a pure reactance over other frequencies. The only advice that can be given to students who have difficulty in understanding the behaviour is that calculations indicate that the networks behave this way, and that practical measurements prove it.

7. ATTENUATION OF T NETWORKS OF PURE REACTANCES.

7.1 The networks dealt with in Fig. 7 have little or no application in these books - they were dealt with because they provided a simple introduction to the slightly more complex behaviour of Fig. 8. The networks of Fig. 8, however, are very important, as they form the basic types of low-pass and high-pass filter sections.

7.2 Whilst the two sections of Fig. 8 are behaving as resistances, they are capable of absorbing power from a generator connected to them, that is, the networks as a whole can take power from the generator. The individual reactances, however, still behave as pure reactances and, therefore, cannot dissipate any of the energy taken by the network as a whole from the generator. All of the power taken from the generator, therefore, is passed on to the termination connected to the other end of the sections. This means that the attenuation over the band of frequencies to which the networks offer a purely resistive characteristic impedance is zero. When the networks become purely reactive, however, they cannot absorb power from a generator connected to them, and the current and voltage will be 90° out of phase at all points. It would be physically possible for such a network to produce attenuation, that is, reduce the amplitudes of the current and voltage as they pass through them, because a decrease in voltage and current does not represent a dissipation of power when the current and voltage are 90° out of phase with each other.

7.3 From the general statement above, it should be clear that the attenuation produced by Figs. 8a and 8b is zero whilst the characteristic impedance is a pure resistance. The fact that attenuation is produced when the networks are reactive can be proved by using equation (3). A section, such as Fig. 8a, produces a phase shift of exactly 180° when the frequency is above the value at which $X_2 = \frac{X_1}{4}$.

8. PROTOTYPE L.P. AND H.P. FILTER SECTIONS.

8.1 Paragraphs 7.2 and 7.3 above indicate that a section of the type shown in Fig. 8a could be used as a low-pass filter because it produces zero attenuation between zero frequency and that at which $X_2 = \frac{X_1}{4}$ and attenuates all frequencies above this. Also, a section of the type shown in Fig. 8b could be used as a high-pass filter, because it produces zero attenuation at all frequencies above that at which $X_2 = \frac{X_1}{4}$ and attenuates all frequencies below this. These sections are called the basic or prototype filter sections because from these are developed the more complicated sections used in practical filters. The two sections, as they appear in Fig. 8, are not satisfactory for two reasons.

8.2 The first reason is that the attenuation does not rise sufficiently sharply beyond the frequency at which $Z_2 = \frac{Z_1}{4}$, called the "cut-off frequency" and designated fc. This

/is

is shown in Fig. 9 for both the L.P. and H.P. filter sections, the attenuation outside the pass band being calculated.

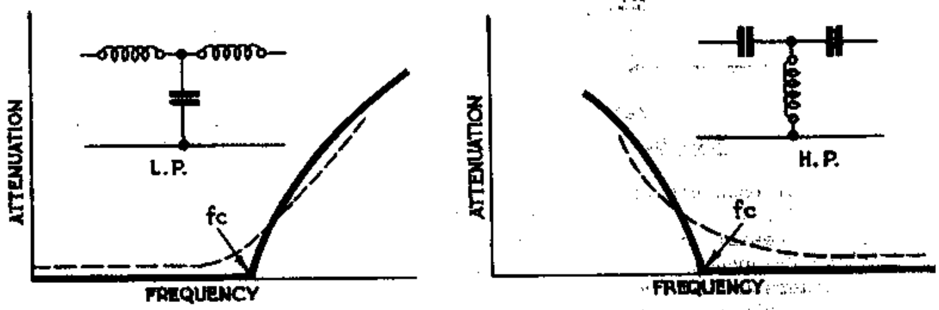


FIG. 9. ATTENUATION VERSUS FREQUENCY CURVES.

When resistance is present, as it always is because it is not possible to wind inductances without some resistance, the position is worsened because the presence of the resistance produces a gradual rather than a sharp cut-off, besides some attenuation in the pass band, as shown by the dotted lines of Fig. 9.

8.3 The second reason is that the characteristic impedance of such sections varies widely over the pass band. This can be shown as follows for the L.P. section.

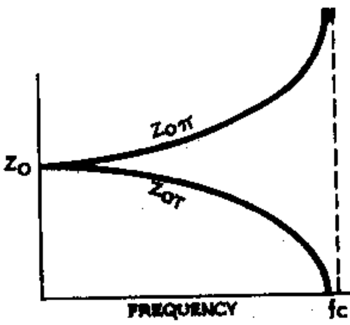
From equation (1) -

$$Z_o = \sqrt{Z_1 Z_2 + \frac{Z_1^2}{4}}$$

can be derived the equation for Fig. 8a -

$$Z_o = \sqrt{\frac{L}{C} - \frac{\omega^2 L^2}{4}} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (7)$$

An examination of equation (7) indicates that, when the frequency is zero, Z_o is $\sqrt{\frac{L}{C}}$ because $\frac{\omega^2 L^2}{4}$ is zero. As the frequency is increased, however, $\frac{\omega^2 L^2}{4}$ increases until, at the cut-off frequency, it equals $\frac{L}{C}$, so that over this band of frequencies, the pass band for Fig. 8a, the characteristic impedance varies between $\sqrt{\frac{L}{C}}$ and zero. This is shown in Fig. 10.



CHARACTERISTIC IMPEDANCE VERSUS FREQUENCY.

FIG. 10.

By the same reasoning, the H.P. section of Fig. 8b will exhibit a characteristic impedance which varies between $\sqrt{\frac{L}{C}}$ at infinite frequency and zero at the cut-off frequency. Such filter sections, therefore, could not be terminated in a single resistance as, if the resistance matched the characteristic impedance of the section at low frequencies, there would be severe reflection at high frequencies, and vice versa.

Included in Fig. 10 is the curve for the characteristic impedance for a π section low-pass filter over its pass band.

9. CUT-OFF FREQUENCIES FOR L.P. AND H.P. FILTERS.

9.1 The cut-off frequency, fc, occurs in both types when $X_2 = \frac{X_1}{4}$, as previously stated.

For the low-pass section -

$$X_1 = \omega L \text{ and } X_2 = \frac{1}{\omega C}$$

$$\therefore \text{ at fc, } \frac{1}{\omega C} = \frac{\omega L}{4}$$

$$\therefore \omega^2 LC = 4$$

$$\therefore \omega^2 = \frac{4}{LC}$$

$$\therefore 4\pi^2 fc^2 = \frac{4}{LC}$$

$$\therefore fc^2 = \frac{4}{4\pi^2 LC}$$

$$\therefore fc = \frac{1}{\pi\sqrt{LC}} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (8)$$

For the high-pass section -

$$X_1 = \frac{1}{\omega C} \text{ and } X_2 = \omega L$$

$$\therefore \text{ at fc, } \omega L = \frac{\frac{1}{\omega C}}{4}$$

$$\therefore \omega L = \frac{1}{4\omega C}$$

$$\therefore \omega^2 = \frac{1}{4LC}$$

$$4\pi^2 fc^2 = \frac{1}{4LC}$$

$$\therefore fc^2 = \frac{1}{16\pi^2 LC}$$

$$\therefore fc = \frac{1}{4\pi\sqrt{LC}} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (9)$$

Equations (8) and (9) are the cut-off frequencies for low-pass and high-pass filter prototype sections in terms of the inductance and capacitance used.

10. BAND-PASS FILTERS.

10.1 A band-pass filter is essentially a low-pass filter in series with a high-pass filter. For example, if it is desired to pass only the band 8-10 kc/s from a band containing all frequencies from zero to infinity, the low-pass filter would have a cut-off frequency of 10 kc/s and the high-pass filter a cut-off frequency of 8 kc/s. The two filters are shown in Fig. 11a and are combined together, as they are in practice, in Fig. 11b.

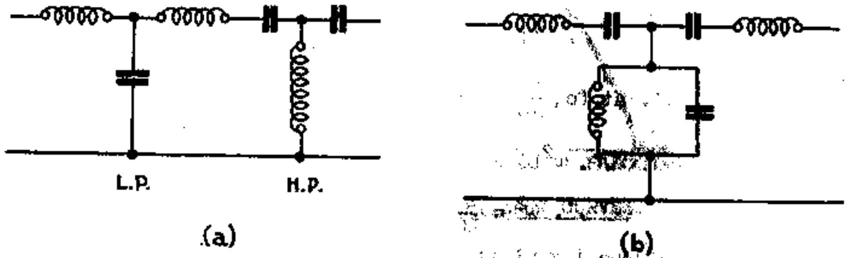


FIG. 11. BAND-PASS FILTER.

10.2 Fig. 12 shows the attenuation versus frequency characteristics of a band-pass filter of the type shown in Fig. 11b. The theoretical curve is shown in full line and the actual values shown dotted.

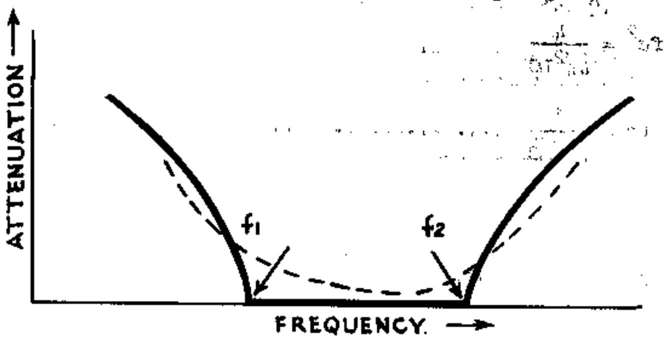


FIG. 12. ATTENUATION VERSUS FREQUENCY CHARACTERISTIC.

10.3 As it is impossible in practice to construct inductances and condensers which are entirely non-resistive, there is always a slight amount of attenuation in the pass-band of a filter. In general, this does not exceed 0.5 db, except in the neighbourhood of the cut-off frequencies, f_1 and f_2, where the effect of the resistance is to cause a rounding-off of the attenuation-frequency characteristic.

11. m-DERIVED FILTER SECTIONS.

11.1 A higher attenuation at frequencies just outside the pass-band, together with a sharper cut-off, can be obtained by the use of "m-derived" filter sections. Fig. 13 shows such a section for a low-pass filter.

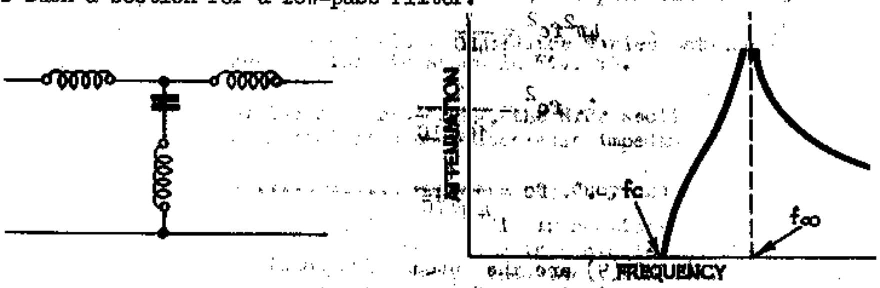


FIG. 13. m-DERIVED L.P. FILTER SECTION.

As shown in Fig. 13a, the shunt element is a series circuit and is designed to be resonant at a frequency just above the cut-off frequency. At this frequency, designated f_∞ in Fig. 13b, the shunt element is a short-circuit and the attenuation is infinite. However, past f_∞ the attenuation falls, so that, in practical filters, a number of such sections is used, each having a successively higher value of f_∞ and so providing the necessary high attenuation throughout the stop band.

/11.2

11.2 These sections are called m-derived sections, because of the relationship between the series elements of the prototype from which they are designed and the series elements in the derived type. The different sections in a filter must have the same characteristic impedance in order to prevent reflection between the sections. If the characteristic impedances are the same at all frequencies, then the different sections will have the same transmission bands as, in this band, and in this band alone, Z_o is a pure resistance. The series elements of a derived type will, of course, differ in value from the prototype because of the inclusion of other components in the shunt element. Thus, in Fig. 14, Z_1 and Z_2 are the series and shunt elements, respectively, of the prototype, and $Z_1{}^1$ and $Z_2{}^1$ are the series and shunt elements, respectively, for a derived type.

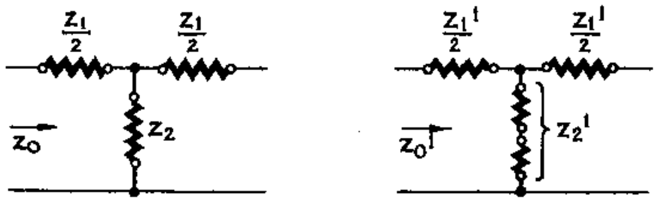


FIG. 14. PROTOTYPE AND m-DERIVED SECTIONS.

As mentioned above, the transmission bands match when the characteristic impedance match, the latter being necessary to prevent reflection. The relation between Z_1 and $Z_1{}^1$ has been designated -

$$Z_1{}^1 = mZ_1$$

$$\text{Also } Z_o{}^1 = Z_o$$

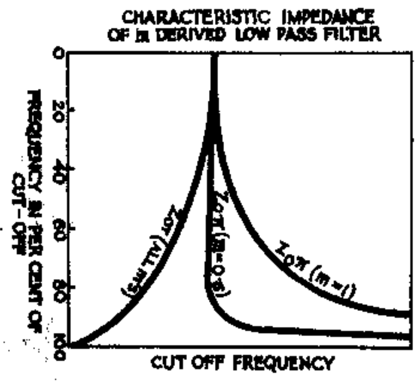
11.3 It is possible to obtain a value for m for any value of fc and f∞ required (with which this Paper is not concerned) and, from a prototype, to design the required m-derived structures. This is a design problem and will not be attempted here.

12. PRACTICAL FILTERS.

12.1 A practical filter consists of prototype sections and m-derived sections designed to produce the required characteristic impedance (which is the impedance between which the filter works), together with a sharp cut-off and a high attenuation in the stop band and a low and even attenuation in the pass band.

12.2 The variation in characteristic impedance over the pass band has now to be corrected. This is done by taking advantage of the fact that the characteristic impedance of a section with an m of 0.6 is constant over practically the whole of the pass band. From equations (4) and (7), it is evident that the characteristic impedance of a prototype section L.P. filter is -

$$\frac{\frac{L}{C}}{\sqrt{\frac{L}{C} - \frac{\omega^2 L^2}{4}}} \quad \text{........................(10)}$$



CHARACTERISTIC IMPEDANCE VERSUS FREQUENCY m-DERIVED SECTIONS.

FIG. 15.

Equation (10) indicates that, at zero frequency, the characteristic impedance of a prototype π section of a low-pass filter is $\sqrt{\frac{L}{C}}$ as in the T prototype, but at the cut-off frequency it is infinity because the denominator of equation (10) is zero at that frequency. By working out the characteristic impedance at all frequencies in the pass band for all values of m for an m-derived π section, that is, by experimental mathematics, it is found that the characteristic impedance of an m-derived π section with an m of 0.6 is substantially constant over the pass band, as shown in Fig. 15.

/12.3

12.3 The procedure now is to design an m-derived T section with an m of 0.6, split the shunt element into two equal parts and, with these two half T sections, terminate the filter with them arranged as half π sections, as in Fig. 16.

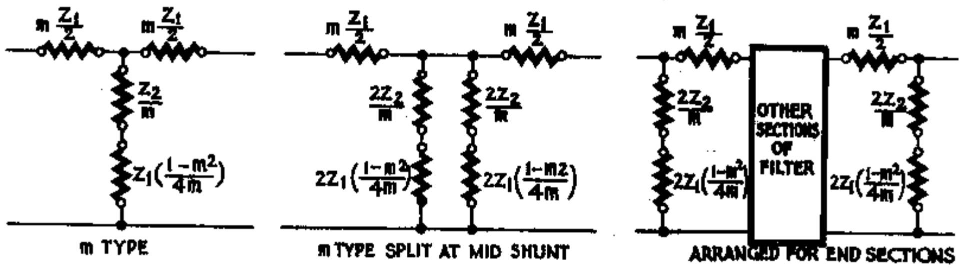
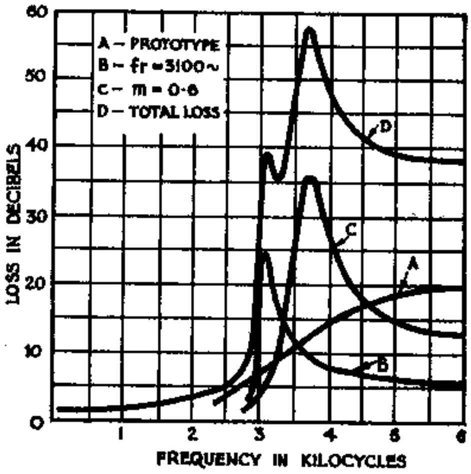


FIG. 16. ARRANGEMENT OF END SECTIONS.

Fig. 17 shows a 3 kilocycle low-pass filter, together with curves giving some idea of the purpose of each unit in a practical filter. Such filters are frequently called Composite Filters.



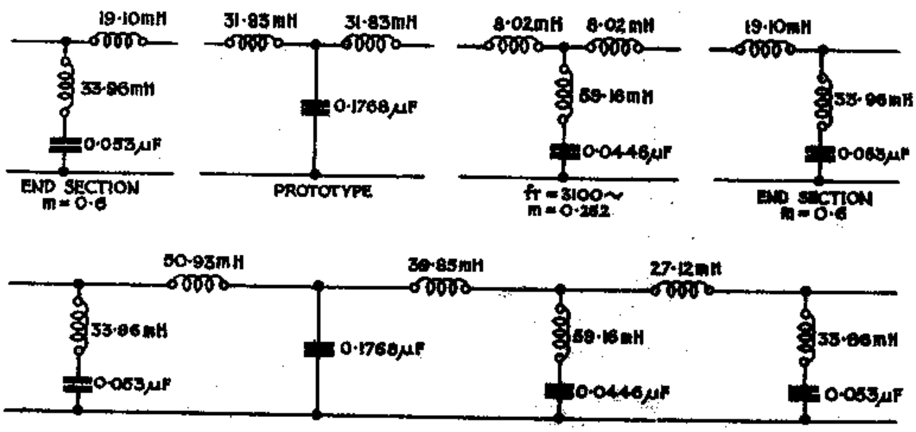


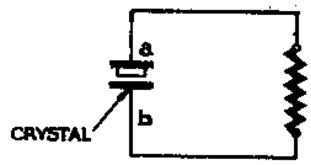
FIG. 17. TYPICAL LOW-PASS FILTER.

13. CRYSTAL FILTERS.

13.1 Certain materials, notably some species of quartz, exhibit the property of developing an e.m.f. across them when subject to pressure or tension. For example, when a mechanical pressure is applied to such a material, an e.m.f. is developed across two opposite faces of the material, which is in crystalline form. When the pressure is removed, the natural elasticity of the material allows it to resume its former state and the direction of the e.m.f. is reversed. This action is reversible in that when an alternating e.m.f. is applied to two opposite faces, slight changes of shape take place which cause the material to vibrate. As with all objects, a piece of such material will exhibit a natural frequency of vibration, that is, a frequency at which the applied alternating e.m.f. will cause the material to execute vibrations at a maximum amplitude.

13.2 As the material vibrates, therefore, alternating voltages are developed across opposite faces, and it would be expected that the larger the amplitude of vibration, the greater will be the amplitude of the voltage developed across it, reaching a maximum at the natural frequency of the specimen being used, and this is found to be so. Specimens of different size display different natural frequencies so that preparation of a specimen to display a particular natural frequency consists in cutting a section of a large specimen to the desired dimensions. In practice, the thickness of the specimen largely determines its natural frequency. The completed specimen is called a "crystal," and preparation of the crystal mainly consists of cutting a specimen from a larger one and grinding it to the size required to produce the required natural frequency of vibration.

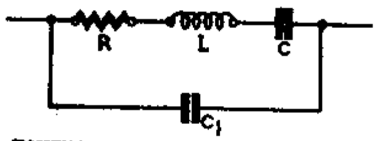
13.3 As discussed above, when a crystal vibrates, an alternating e.m.f. is developed across two of its opposite faces. The action is perhaps best explained electronically from Fig. 18. The crystal in Fig. 18 is caused to vibrate by means not shown, the opposite faces (across which the e.m.f. is developed) being connected to some circuit. During the half-cycle when side a is positive to side b, side b will exhibit a surplus of electrons that were transferred from side a (which exhibits a deficit) via the circuit connected to the crystal. During the other half-cycle, side a is negative and, therefore, displays a surplus of electrons collected from side b via the circuit. Thus, the alternating voltage developed across the crystal faces causes an alternating current to flow through the circuit. This alternating current will be a maximum at the resonant frequency of the crystal, that is, when the amplitude of vibration of the crystal and, therefore, the voltage developed across it is a maximum.



VIBRATING CRYSTAL IN CIRCUIT.

FIG. 18.

13.4 In this respect, the crystal behaves as a series resonant circuit, the equivalent circuit being shown in Fig. 19. L is the effective mechanical inductance due to the mass of the crystal, R is the effective resistance due to losses produced by intermolecular friction as it vibrates, and C is the effective mechanical capacity due to the elasticity of the crystal. C_1 is the capacity between the faces of the crystal due to its straight condenser action.

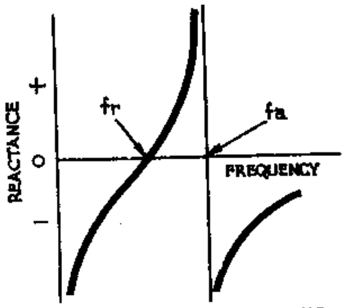


EQUIVALENT CIRCUIT OF CRYSTAL.

FIG. 19.

/13.5

13.5 The reactance curve of a crystal is shown in Fig. 20. From Fig. 20 it will be seen that the crystal has a resonant frequency fr and an anti-resonant frequency fa. The resonant frequency is due to the series circuit alone, whilst the anti-resonant frequency is due to the shunting effect of C_1 in Fig. 19. The ratio of C_1 to C is a constant for any given material; for quartz it is 125 to 1. This gives an anti-resonant frequency which is 0.4% higher than the resonant frequency for quartz.

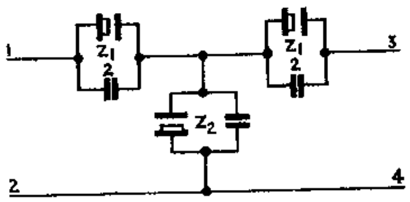


REACTANCE CURVES OF CRYSTAL.

FIG. 20.

13.6 The desirable feature of crystals is their low equivalent resistance, being near enough to zero for most practical purposes. This makes crystals ideal for such uses as sharply tuned circuits and as reactive elements in filters where a sharp cut-off is required.

13.7 Fig. 21 shows a T section crystal filter with its associated reactance curve and attenuation curve.



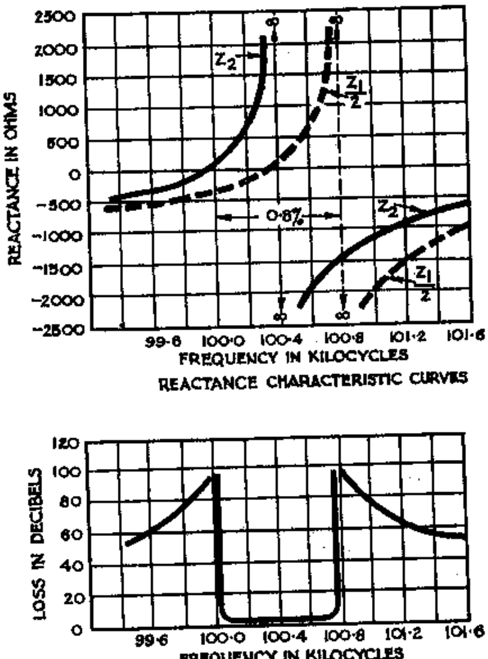


FIG. 21. T FILTER USING CRYSTALS.

The crystal elements are so selected that the resonant frequency of the series arm (zero reactance) coincides with the anti-resonant frequency of the shunt arm (infinite reactance). Under this condition, the network will have, from merely a general examination, maximum attenuation at the anti-resonant frequency of the series arm (infinite series reactance) and at the resonant frequency of the shunt arm (zero shunt reactance).

As in the preceding filters, the pass band will extend over the frequency band in which Z_1 and $Z_2 + \frac{Z_1}{4}$ are opposite types of reactance. The curve for $\frac{Z_1}{4} + Z_2$ can be obtained by adding half the values of the $\frac{Z_1}{2}$ curve of Fig. 21 to the Z_2 curve. The resultant curve will exhibit much the same resonant and the same anti-resonant frequencies as Z_2, and will exhibit a negative reactance below and a positive reactance above resonance. Similarly, the curve for Z_1 can be obtained by doubling all values of the $\frac{Z_1}{2}$ curve in Fig. 21. Such a curve will display the same resonant and anti-resonant frequencies as the $\frac{Z_1}{2}$ curve.

Thus, the pass band is near enough to between the resonant frequency of Z_2 and the anti-resonant frequency of $\frac{Z_1}{2}$. This means that the pass band of such a filter is only 0.8% of the mid-frequency, for example, in Fig. 21 the pass band is only 800 c/s at 100 kc/s, which is too narrow for a voice frequency channel. Where only a very narrow band is desired (much narrower than the 125 to 1 ratio of C_1 to C can provide), the ratio is increased by a condenser shunting the crystal, as in Fig. 21, which /effectively

effectively reduces the band-width passed.

13.8 The crystal filters used are usually of "lattice" formation, those dealt with up to the present being of "ladder" formation. (When T or π sections are connected one after the other, the structure is not unlike a ladder.) Fig. 22 shows a lattice filter, together with its reactance and attenuation curves.

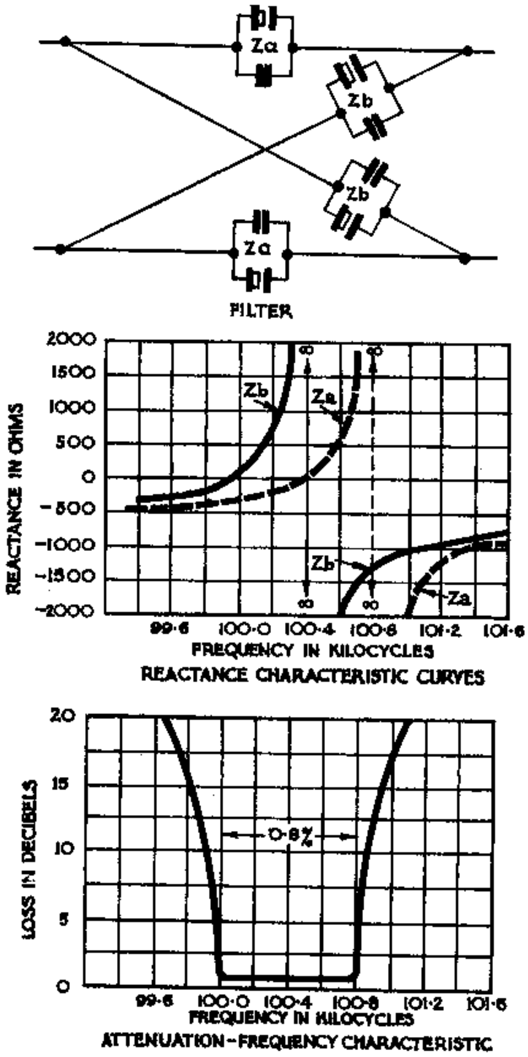


FIG. 22. CHARACTERISTICS OF LATTICE FILTER USING CRYSTALS.

The series arms are alike, as are the two shunt arms, but the series and shunt arms differ, as will be seen from the reactance curves. The pass band extends over those frequencies at which the series and shunt arms are opposite types of reactance. Again, as in Fig. 21, the pass band is only 0.8% of the mid-frequency.

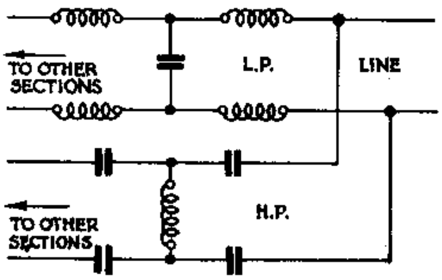
To produce the wider pass bands necessary in practice, an inductance is used in series with the crystals. Within certain limits, this series inductance does not introduce enough resistance to interfere with the sharpness of cut-off produced by the crystal, and, by a suitable adjustment of the shunt condenser across the crystals, pass bands of any desired width may be obtained within the limits permitted by the inductance.

14. PARALLEL CONNECTION OF FILTERS.

14.1 It is frequently necessary to operate filters in parallel. In Fig. 13 of Paper No. 2, there are three examples of this practice. The L.P. and H.P. line filters are parallel connected on their line sides, the two directional filters are parallel connected on the H.P. line filter sides and the two groups of three band-pass filters in the channel circuits are parallel connected on the side of the appropriate amplifier.

14.2 Generally speaking, all filters have to work into a circuit that displays an impedance which is independent of frequency and, therefore, is resistive. As an example, an aerial line into which the line filter group of Fig. 13 in Paper No. 2 works is usually, for all practical purposes, near enough to a zero angle line. This means that the impedance looking into the line filter group from the line, that is, the impedance of the H.P. and L.P. line filters in parallel, should be a constant resistance.

14.3 As discussed previously, a constant resistance termination is provided for a filter by designing a derived section with an m of 0.6, splitting this section down its shunt element and terminating each end of the filter with the half sections so formed arranged as half sections. This effect is achieved in the parallel connection by using one of the T sections, either prototype or derived, in each filter to terminate the filters on the sides which are to be paralleled, as in Fig. 23. From Fig. 23 it will be seen that the first series capacity and shunt inductance of the H.P. filter provide a shunt path containing capacitance and inductance in series across the input to the filter and are equivalent, therefore, to the shunt element of an m-derived L.P. section. This means that merely connecting the two filters in parallel, as described above, provides an m-derived termination for the group. Where the components employed in the terminating sections do not provide an m of 0.6, an adjustment can be made to the first series inductance of the L.P. section and the first series capacitance of the H.P. section to achieve this.



L.P. AND H.P. FILTERS IN PARALLEL.

FIG. 23.

14.4 When paralleling a number of band-pass filters, it is necessary to ensure that the impedance of each filter rises sharply outside its pass band. This is done by using a suitable design of T section to terminate each filter at the end to be paralleled. By this means, each filter in its pass band will be shunted by the high impedance of the remaining paralleled filters outside their pass bands, and the characteristics of each filter in its pass band will be unaffected by the presence of the others. Where the pass bands are not spaced widely apart, it is necessary to go a step further, and this is generally the case. As before, the filters are connected in parallel, but there will now be a difference in that the impedances shunted across each filter by its neighbours will be lower owing to the narrower gaps between the pass bands.

The filters above and below will contribute reactive components which are opposite in sign and, therefore, will tend to correct one another. For example, above the pass band of filter 1 in Fig. 24, the reactive component is inductive, whilst below the pass band of filter 2 the reactive component is capacitive. Even though these reactances be low in value they will correct one another, as will the inductive reactance above the pass band of filter 2 and the capacitive reactance below the pass band of filter 3.

In Fig. 24, however, there is no compensating inductive reactance to correct the low capacitive reactance immediately below the pass band of filter 1, and no

/compensating

compensating capacitive reactance to correct the low inductive reactance immediately above the pass band of filter 3. It is often necessary, therefore, to connect an auxiliary network, called a "compensating network," across the whole filter group to correct these residual low reactances.

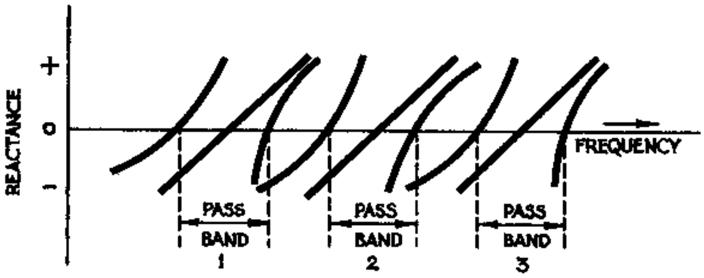


FIG. 24. B.P. FILTERS IN PARALLEL.

15. EQUALISERS.

15.1 As mentioned in Paper No. 1, one of the factors which tend to decrease the intelligibility of telephone speech is unequal attenuation of the currents of different frequencies as they are transmitted over the circuits. For example, the attenuation of an open wire circuit is greater for the higher frequencies than for the lower frequencies, and this difference in attenuation is directly proportional to the length of line. Therefore, when long circuits are employed, it is frequently necessary to employ attenuation equalisers to correct the unequal attenuation of the line. These equalisers are usually associated with the repeaters or amplifiers which are included in the circuit at various points, in order to provide a uniform level of current at the receiving end over the frequency range employed. The unequal attenuation of different frequencies is even more marked in unloaded cable circuits.

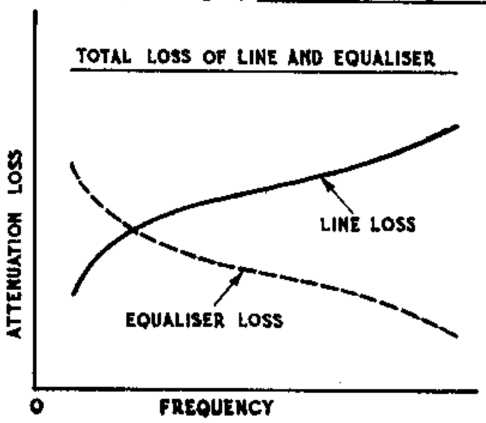


FIG. 25. PRINCIPLE OF EQUALISER.

15.2 Attenuation equalisers are networks consisting of inductances, condensers and resistances which are so proportioned and arranged that their attenuation-frequency characteristics are complementary to the line characteristics that produce the distortion. In brief, the total loss of the line plus the loss produced by the equaliser is the same for all frequencies within the band of frequencies concerned. This is shown in Fig. 25.

15.3 One of the simplest types of equalisers, shown schematically in Fig. 26, is bridged directly across the circuit to be corrected. The impedance of such a bridged equaliser must be low enough at certain frequencies to allow sufficient current to flow through it to produce the required losses at these frequencies. Accordingly, the equaliser circuit naturally changes the circuit impedance, particularly at the frequencies where the equaliser is to provide a substantial loss. This introduces an impedance irregularity of a sizeable value into the circuit. The use of bridged equalisers of this type, therefore, has definite limitations in practice.

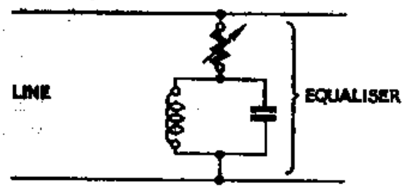


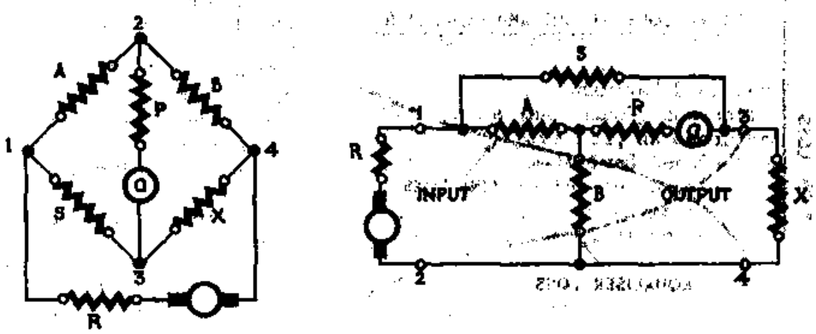
FIG. 26. SIMPLE BRIDGED EQUALISER.

/15.4

15.4 In long circuits equipped with telephone repeaters, the desired equalising effects can be obtained without introducing an appreciable impedance irregularity by inserting equalising networks at the mid-point of the primary sides of the repeater input transformers. Instead of changing the net loss of the line, however, this arrangement changes the overall gain-frequency characteristic of the repeater to match reasonably closely the loss-frequency characteristic of the line. In other words, for the frequencies where the line loss is high the repeater gain is also high, and vice versa. The overall loss-frequency characteristic of the line and repeater together is then reasonably uniform over the transmitted frequency band.

15.5 Bridged T Equaliser. Both of the above methods of equalisation give satisfactory results where the amount of attenuation distortion to be corrected is relatively small. The use of either of these methods to correct a large attenuation distortion might result in an impedance irregularity of such a magnitude as to more than offset the benefits obtained by equalising. To equalise for these relatively large amounts of attenuation distortion, a somewhat more complex equalising network, in the form of a Bridged T structure, may be used. This equaliser is designed to have a constant impedance over the entire frequency band transmitted.

As the name implies, the Bridged T equaliser is built in the general form of a T network, but it has an additional impedance path bridged across the series elements. This latter path controls the loss of the equaliser. The elements of the Bridged T equaliser are connected in a Wheatstone bridge arrangement, and the principle of its operation may be best grasped by first referring to the ordinary Wheatstone bridge circuit shown in Fig. 27a.



(a) Wheatstone Bridge Circuit.

(b) Wheatstone Bridge in the Form of a "T" Network.

FIG. 27. BRIDGE PRINCIPLE APPLIED TO BRIDGED T EQUALISER.

Now rearrange this bridge circuit in the form of a T network where the series elements are bridged by the impedance S, as shown in Fig. 27b. The T network proper is formed by A, P and B with S as the bridging impedance, while R and X now become the input and output impedances, respectively. Next, change the impedances R, A, P and X to resistances of equal value, which may then all be designated as R. For reasons to be explained later, the impedance S and B will also be redesignated as Z_{11} and Z_{21}, respectively. Then, as illustrated in Fig. 28, the bridge which was balanced when

$$SB = XA \text{ is now balanced when -}$$

$$Z_{11}Z_{21} = R \times R \text{ or } R^2.$$

/When

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COURSE OF TECHNICAL INSTRUCTION.

VOICE FREQUENCY REPEATERS.

CONTENTS.

1. INTRODUCTION.

1.1 Prior to the invention of the thermionic valve and the consequent development of the voice frequency (V.F.) repeater, the most serious hindrance to the extension of trunk line communication was the excessive attenuation introduced by the long trunk lines. This could be partially overcome only by the costly expedients of "loading" or by using heavy gauge copper conductors, and a number of 600 lb. aerial trunk lines were installed for this reason. With the introduction of the V.F. repeater, this limitation was removed and it is now possible, by installing repeaters at regular intervals along the route, to extend trunk line communication over very long distances without departing from the requirements of modern transmission standards.

1.2 As previously referred to in Long Line Equipment I, the Overall Transmission Performance Standard requires that when any two subscribers connected to two different telephone exchanges within the Commonwealth network are interconnected via junction and trunk lines, the maximum permissible attenuation between those two terminal exchanges (at a frequency of 1 kc/s) is limited to 15 db. To ensure that this figure is not exceeded, it is necessary, in practice, to limit the attenuation of certain types of trunk lines to 6 db; and, furthermore, to ensure that trunk lines between Main, Primary, and Secondary Trunk Centres are zero loss equivalent.

1.3 Because of line attenuation, however, the distance over which V.F. transmission of the required overall equivalent may be obtained, is restricted. For example, the attenuation of a 200 lb. H.D.C. aerial trunk line at a frequency of 1 kc/s is approximately 0.06 db/mile, resulting in a limiting length of about 100 miles for a permissible transmission loss of 6 db. In order to obtain transmission within the required standards and over greater distances, it is necessary to use either heavier gauge line conductors or thermionic valve amplifier units, known as V.F. repeaters. For economic reasons the latter method is preferred.

1.4 Thermionic valve amplifiers are unidirectional in operation in that, whilst a signal voltage applied to the input terminals (that is, between grid and cathode) is amplified in the output circuit (that is, between anode and cathode), a signal voltage applied to the output terminals is not of amplified form in the input circuit. When operating as a V.F. repeater, however, the thermionic valve must be capable of amplifying the powers transmitted in each direction over the trunk line and this necessity, together with other considerations referred to later, has led to the development of three general types of V.F. repeaters, namely -

(i) type 21 (2-wire - one element),
(ii) type 22 (2-wire - two element) and
(iii) the 4-wire type.

Two-wire repeaters are commonly used in aerial trunk line circuits, and 4-wire repeaters in trunk cable circuits.

1.5 This Paper deals with the operation and provision of V.F. repeaters on aerial and cable trunk line circuits.

2. TYPE 21 V.F. REPEATER.

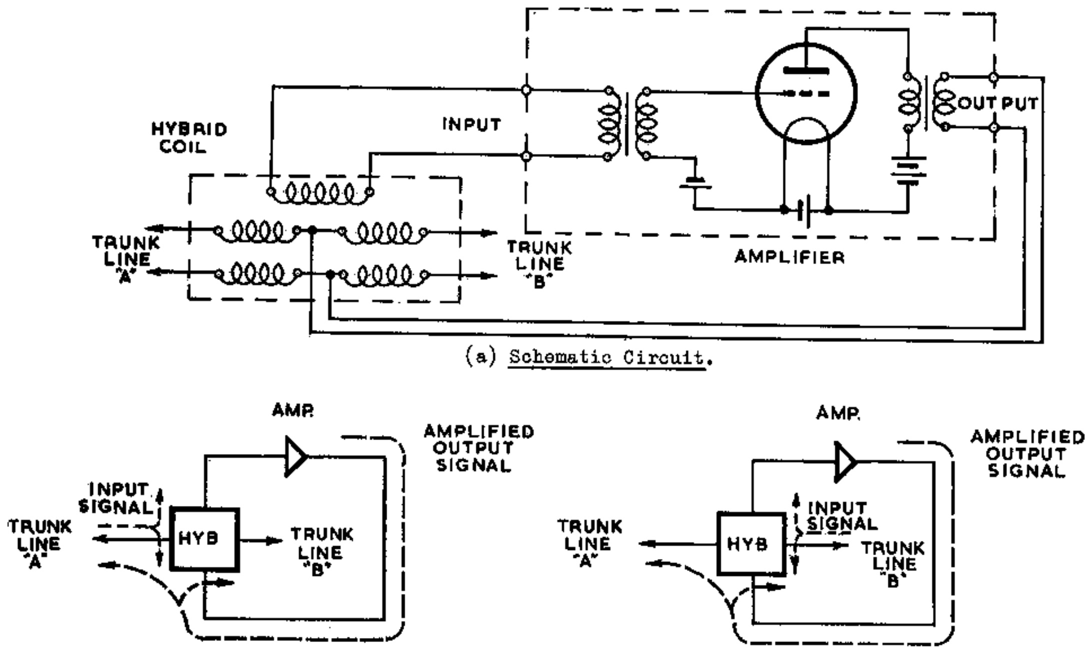
2.1 The type 21 V.F. repeater, although now obsolete, is briefly mentioned here, as it represents a stage in the development of the type 22 V.F. repeater in general use.

2.2 The repeater (see Fig. 1) consists essentially of a single thermionic valve amplifier stage with input and output connected to a special three winding transformer known as a hybrid coil or differential transformer. V.F. signals arriving from either line are applied via the hybrid coil to the input and output circuits of the amplifier. The portion entering the input circuit is amplified by the valve and applied to the midpoint of the line windings of the hybrid coil. Here it divides, half returning back along the trunk line from which the original signal was received and serving no useful purpose, the other half being applied to the distant trunk line section.

2.3 An essential condition for the satisfactory operation of the repeater is that the impedances of the trunk line sections on either side of the hybrid coil must be identical over the range of voice frequencies transmitted over the circuit. When the impedances of the lines are unequal, portion of the amplified signal voltage applied to the midpoint of the line windings of the hybrid coil is reapplied back to the input of the amplifier and further amplified. This process continues in a cumulative manner until the amplifier commences to oscillate or "sing."

2.4 Although this form of repeater has the economic advantage of using only one thermionic valve for amplifying both directions of transmission, its application is limited because of the practical difficulty of maintaining a close balance between the trunk line impedances. This disadvantage led to the development of the type 22 V.F. repeater for use on 2-wire trunk line circuits.

/ Fig. 1.



(a) Schematic Circuit.

(b) Equivalent Block Schematic Circuit.

V.F. REPEATER - TYPE 21.

FIG. 1.

3. TYPE 22 V.F. REPEATER.

3.1 In the type 22 V.F. repeater two amplifier units are provided, one for each direction of transmission. The basic arrangement is shown in Fig. 2.

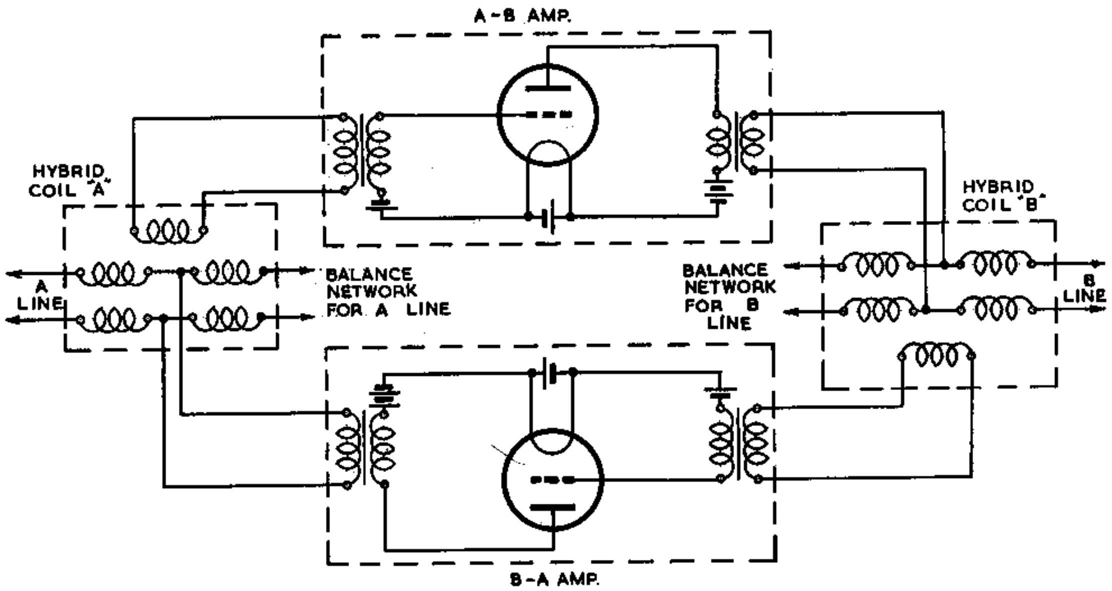
3.2 To prevent the possibility of "singing" with this arrangement, it is necessary to ensure that the amplified signals in the output of either amplifier are not applied to the input of the other amplifier. This necessitates the separation of the opposite directions of transmission, which can be done in a V.F. repeater only by the use of hybrid coils - filters are not satisfactory - because the same frequency range is transmitted in opposite directions.

To ensure that the hybrid coils isolate the input of each amplifier from the output of the other, the impedance of each balance network must simulate the impedance of the trunk line with which it is associated, over the range of voice frequencies transmitted over the circuit.

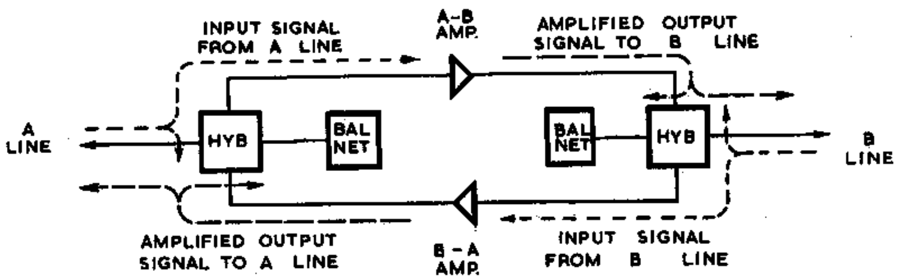
Fig. 2a shows a typical arrangement of the hybrid coil connections. In some repeaters the input and output connections of the amplifier to the hybrid coil are reversed.

/ 3.3

3.3 With either type of hybrid connection, the V.F. signal from trunk line A is applied to hybrid coil "A", where it divides, half being dissipated in the output impedance of the B-A amplifier and half entering the input circuit of the A-B amplifier. The amplified output from the A-B amplifier is applied to the midpoint of the line windings of hybrid coil "B" where it divides, half being dissipated in the balance network, and half passing to trunk line B. Transmission in the reverse direction is similar, the signals in this case being amplified by the B-A amplifier.



(a) Schematic Circuit.



(b) Equivalent Block Schematic Circuit.

V.F. REPEATER - TYPE 22.

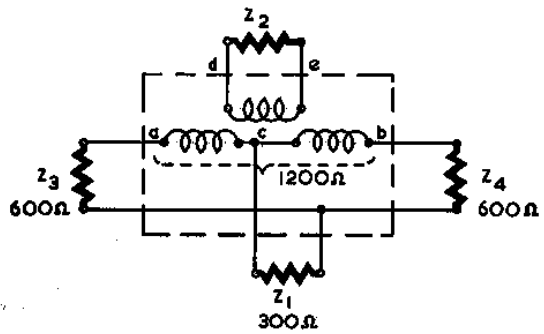
FIG. 2.

4. HYBRID COIL - THEORY OF OPERATION.

4.1 In this section, the operation of the hybrid coil is considered in more detail for signals passing through it in the various directions of transmission.

4.2 The hybrid coil may be of the unbalanced or balanced type. In the unbalanced type, which is the simpler arrangement, there are windings in one side of the line only. This type is used in such cases as submarine cable operation. In the balanced type, which is the type more commonly used in practice, these windings are divided into two halves, one half of the winding being in each side of the line to preserve balanced impedance conditions for crosstalk considerations. The theory of operation is similar for both types, but for simplicity of explanation, the description is given here for the unbalanced type.

4.3 Impedance Relations of Hybrid Coil. Fig. 3 shows a simplified circuit for an unbalanced hybrid coil in which the amplifier connections, trunk line and balance network have been represented by appropriate impedances. For the satisfactory operation of the circuit, certain impedance conditions must exist. The relative values of these are shown for a trunk line impedance of 600 ohms.



IMPEDANCE RELATIONS OF HYBRID COIL.

FIG. 3.

Notes on Fig. 3 -

Winding a-b is centre tapped at c.

Z1 and Z2 = Input and Output Impedances of Amplifiers.

Z3 = Trunk Line Impedance.

Z4 = Balance Network Impedance.

The impedance of 1,200 ohms shown across the a-b winding represents the impedance (Zr) "reflected" from Z2 when current flows through the a-b winding, and is determined by the formula -

$$Zr = T^2(Z2).$$

where T = turns ratio of $\frac{a-b}{d-e}$ windings.

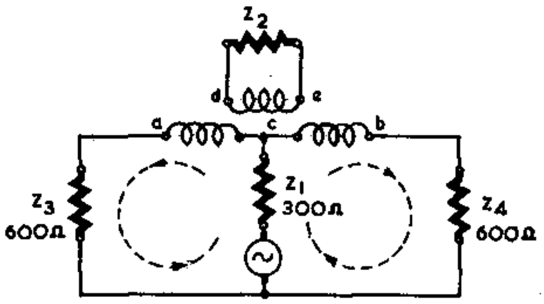
In connection with the effective impedance of the hybrid coil windings however, it is important to note the following -

(i) When the currents flowing in the a-c and b-c windings are equal and in opposite directions, no resultant flux is produced and the windings do not possess impedance, but present only their low ohmic resistance to the flow of an A.C. through them.

(ii) When a current flows in Z2 due to an e.m.f. induced in the d-e winding from the a-c (or b-c) winding only, then the impedance which Z2 "reflects" into the a-c (or b-c) winding is 300 ohms, because halving the effective turns on a transformer winding reduce the "reflected" impedance to one quarter.

/ 4.4

4.4 Transmission from Amplifier Output (Z1) to Trunk Line. Assume an alternating signal voltage applied to the hybrid coil from the output of the amplifier represented by Z1 in Fig. 4. Because Z3 = Z4 and the winding a-b is centre tapped, the resulting signal current divides equally between the trunk line and balance network. The directions of the currents flowing in the hybrid coil windings are those existing at some instant. No resultant flux is produced by the currents flowing through the a-c and b-c windings, and, therefore, no resultant voltage is induced in the d-e winding or applied to the input of the amplifier represented by Z2. The power from the output of the amplifier therefore divides equally between the trunk line and the balance network, no power being applied to the input of the amplifier serving the opposite direction of transmission.

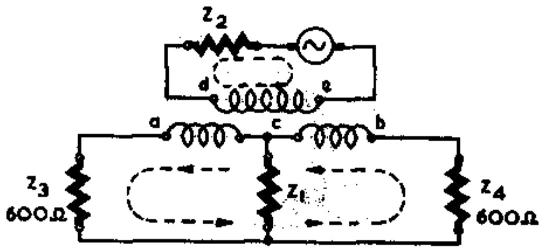


TRANSMISSION FROM Z1 TO TRUNK LINE.

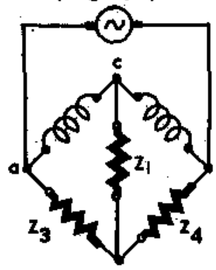
FIG. 4.

Only half the power supplied from the output of the amplifier is applied to the trunk line. A power loss of one half is equivalent to a loss of 3 db, and so the hybrid coil at the output of the amplifier produces an equivalent loss of 3 db in the desired direction of transmission. To ensure conditions of maximum power transfer, the impedance of the supply source Z1 (300 ohms) is correctly "matched" to its load comprising Z3 (600 ohms) and Z4 (600 ohms) in parallel.

4.5 Transmission from Amplifier Output (Z2) to Trunk Line. In some applications of the hybrid coil, the input and output connections of the amplifiers are reversed and Z2 represents the output of the amplifier applying an alternating signal voltage to the d-e winding of the hybrid coil (Fig. 5a). The induced e.m.f. in the a-b winding produces a current flow through Z3 and Z4 in series. No voltage is applied to the input of the amplifier represented by Z1 as it is connected across points of equal potential. This is shown in the equivalent bridge circuit (Fig. 5b).



(a)



(b)

Z3 = Z4.

Winding a-b is centre tapped at c.

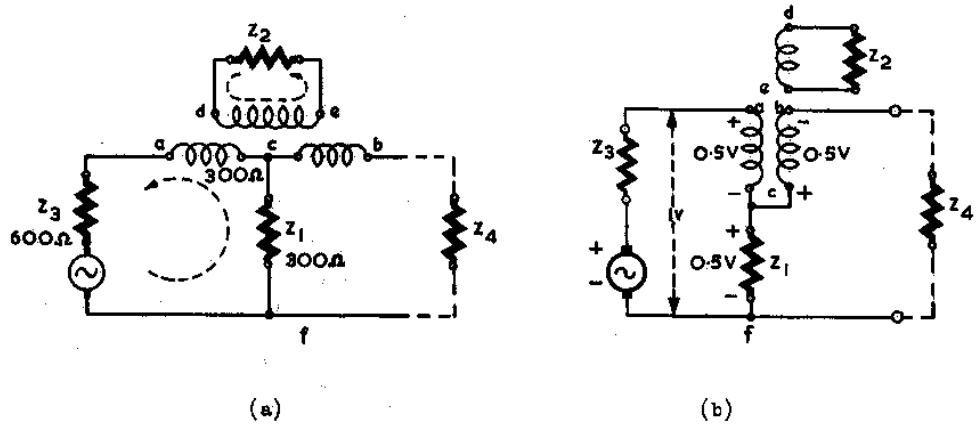
TRANSMISSION FROM Z2 TO TRUNK LINE.

FIG. 5.

As in the previous case, therefore, the power from the output of the amplifier divides equally between the trunk line and the balance network, no power being applied to the input of the amplifier serving the opposite direction; and the hybrid coil again produces an equivalent loss of 3 db in the desired direction of transmission. By suitable selection of the turns ratio of the coil windings, the impedance of the supply source (Z2) "reflects" an impedance of 1,200 ohms across the a-b winding and, therefore, is correctly "matched" to its load comprising Z3 (600 ohms) and Z4 (600 ohms) in series.

/ 4.6

4.6 Transmission from Trunk Line (Z3) to Amplifier Input. Assume now, an alternating signal voltage applied from the trunk line (Z3) to the hybrid coil with the balance network (Z4) initially disconnected as shown in Fig. 6a.



TRANSMISSION FROM TRUNK LINE TO AMPLIFIER INPUT.

FIG. 6.

Under these conditions the resulting signal current flows through Z1 and the a-c winding of the hybrid coil in series. The flux produced by this winding induces a signal voltage across the d-e winding which is applied to Z2. The power from the trunk line, therefore, divides between Z1 and Z2. For maximum power transfer, the turns ratio of the hybrid coil windings must be such that Z2 "reflects" an impedance of 300 ohms into the a-c winding. This, in series with the 300 ohms impedance of Z1 provides the required termination of 600 ohms for the trunk line.

When a voltage is applied to the hybrid coil, therefore, half of that voltage is dropped across Z1, the other half being dropped across the a-c winding of the hybrid coil. This condition is shown in Fig. 6b at some particular instant, for an applied signal of 1 volt.

The voltage across the a-c winding is a voltage of self-induction, this being produced by the current flowing through the a-c winding via Z1. Voltages of mutual induction are induced across the d-e winding and the b-c winding, that across the former being responsible for the loss of half the input power. The direction of the voltage across the b-c winding is indicated and is in the same direction as that across the a-c winding, because the same flux produces both voltages and the two windings are wound in series. The voltage across the b-c winding equals that across the a-c winding, as the two windings have an equal number of turns. Points b and f therefore, are at the same potential, and if the balance network is connected across these points, as it is in practice, no current flows through that network as the necessary condition for a current flow is a difference of potential.

Thus, when power from a trunk line is applied to a V.F. repeater, one half of that power is applied to the input of the appropriate amplifier, the other half being dissipated in the output of the other amplifier. The hybrid coil again produces an equivalent loss of 3 db in the desired direction of transmission. None of the power from the line, however, appears in the balance network, resulting effectively in an infinite loss in this direction of transmission.

/ 4.7

4.7 Balanced Hybrid Coil. In the balanced hybrid coil, a schematic circuit of which is shown in Fig. 7a, the various windings are wound on the same core. The coils 1, 2, 3 and 4 are connected series-aiding and are equal in all respects, having the same number of turns, resistance, self-capacitance, etc. Coil 5 need not have the same number of turns, etc. as coils 1-4, and may provide a step-up or step-down action by suitably choosing the turns ratio. For correct impedance matching, assuming a trunk line impedance (Z3) of 600 ohms, the turns ratio is adjusted so that Z2 "reflects" an impedance of 1,200 ohms across the coils 1-4 when connected series-aiding.

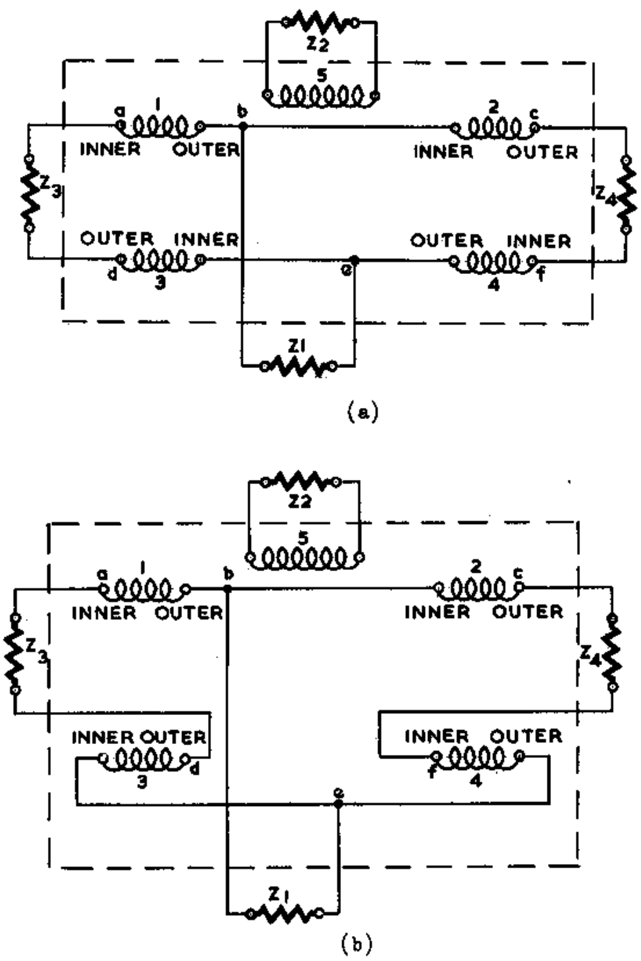


FIG. 7. BALANCED HYBRID COIL.

Fig. 7a is purely a schematic representation of the hybrid coil and is not intended to indicate that the points b and e are the centre points of single windings a-c and d-f respectively. Usually the coils 1, 2, 3 and 4 are wound simultaneously, four spools of wire being used. There are, therefore, four "inner" and four "outer" connections, and care must be taken to interconnect these leads in correct phase relationship. A number of different arrangements are used to connect these coils, a typical arrangement being shown in Fig. 7b. In Fig. 7b, the points have been marked to correspond with the previous figure.

The operation of the balanced hybrid coil is similar to the unbalanced type discussed previously, and may be followed from Fig. 7b for the various directions of transmission.

4.8 Power Loss in Hybrid. Irrespective of the connection to which the input power is applied, the hybrid produces a minimum loss of 3 db in the desired direction of transmission. This condition is summarised in Fig. 8 in which the hybrid coil is replaced by a square in which are indicated the transmission paths existing under perfectly balanced conditions. Between each two adjacent impedances, the equivalent loss of the transmission path is 3 db, whilst between opposite impedances the transmission loss is infinite.

Because of the 3 db loss in each hybrid the gain introduced by a type 22 V.F. repeater in either direction of transmission is 6 db less than the gain of the amplifier appropriate to that direction.

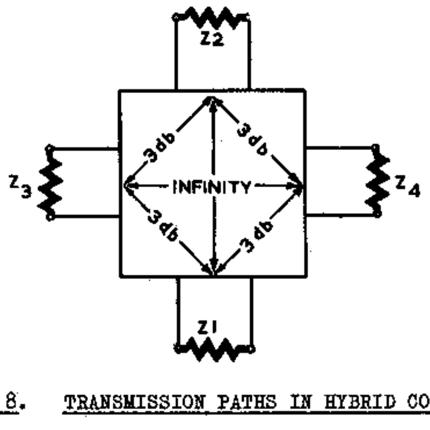


FIG. 8. TRANSMISSION PATHS IN HYBRID COIL.

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