Open-Wire Crosstalk *

By A. G. CHAPMAN

Introduction

THE tendency of communication circuits to crosstalk from one to another was greatly increased by the advent of telephone repeaters and carrier current methods. Telephone repeaters multiplied circuit lengths many times, increased the power applied to the wires, and at the same time made the circuits much more efficient in transmitting crosstalk currents as well as the wanted currents. Carrier current methods added higher ranges of frequency with consequently increased crosstalk coupling. Program transmission service added to the difficulties since circuits for transmitting programs to broadcasting stations must accommodate frequency and volume ranges greater than those required for message telephone circuits.

As these new types of circuits were developed, their application to existing open-wire lines was attended with considerable difficulty from the crosstalk standpoint. Severe restrictions had to be placed on the allocation of pairs of wires for different services in order to keep the crosstalk within tolerable bounds. In many cases the existing lines were retransposed but, nevertheless, there were still important restrictions. While great reduction in crosstalk was obtained by the transposition arrangements the crosstalk reduction was finally limited by unavoidable irregularities in the spacing of the transposition poles and in the spacing of the wires, including differences in wire sag. To further improve matters it was, therefore, necessary to alter the wire configurations so as to reduce the coupling per unit length between the various circuits.

Recently this study of wire configurations has resulted in extensive use of new configurations of open-wire lines in which the two wires of a pair are placed eight inches apart instead of 12 inches, the horizontal separation between wires of different pairs being correspondingly increased. With these eight-inch pairs it has usually been found desirable to discard the time-honored phantoming method of obtaining

* This paper gives a comprehensive discussion of the fundamental principles of crosstalk between open-wire circuits and their application to the transposition design theory and technique which have been developed over a period of years. In this issue of the *Technical Journal* the first half of the paper is published. In the April 1934 issue will be the concluding part, together with an appendix entitled "Calculation of Crosstalk Coefficients."

additional circuits so as to make it possible to obtain a greater number of circuits by more intensive application of carrier current methods.

It is the object of this paper to outline the fundamental principles concerning crosstalk between open-wire circuits and recent developments in transposition design theory and technique which have led to the latest pole line configurations and transposition designs.

To those generally interested in electrical matters it is hoped that this paper will give an insight into the problem of keeping crosstalk in open-wire lines within proper bounds. To those interested in crosstalk it is hoped that the paper will give a useful review of the whole matter and perhaps an insight into the importance of some phenomena which do not seem to be generally appreciated.

The principles set forth in this paper will also be found of considerable interest in connection with problems of control of cable crosstalk, particularly for the high frequencies involved in carrier transmission. It will also be recognized that use is made here of the same general principles as are used in the calculation of effects of impedance irregularities and echoes on repeater operation. These general principles have also been found useful in the development of combinations or arrays of radio antennas of the long horizontal wire type.

The art of crosstalk control in open-wire lines has grown up as a result of the efforts of many workers. The individual contributions are so numerous that it has not been considered practicable in this paper to make individual mention of them except in a few special cases.

General

In the evolution of a satisfactory transposition design technique, complicated electrical actions must be considered and it has been convenient to divide the total crosstalk coupling into various types, all of which may contribute in producing crosstalk between any two circuits in proximity. The first portion of this paper is therefore devoted largely to an examination of the underlying principles and the definition of some of the special terms employed, such as *transverse crosstalk*, *interaction crosstalk*, *reflection crosstalk*, etc. The paper then considers the general effect of transpositions in reducing crosstalk and how this effect depends on the attenuation and phase change accompanying the transmission of communication currents. Consideration is next given to the practical significance of and methods for determining the *crosstalk coefficients* which are used in calculating the crosstalk in a short part of a parallel between two currents. The matter of *type unbalances* inherent in different arrangements of transpositions and used in working from short lengths to long lengths is discussed at length. The next section of the paper is devoted to the effect of constructional irregularities caused by pole spacing, wire sag, "drop bracket" transpositions, etc. Various "non-inductive" wire arrangements are considered. The paper closes with a general discussion of practical transposition design methods based on the principles previously disclosed.

Underlying Principles

The discussion under this heading will cover the general causes of crosstalk coupling between open-wire circuits and the general types into which it is convenient to divide the crosstalk effect. The usual measures of crosstalk coupling will also be discussed.

Causes and Types of Crosstalk

The crosstalk coupling between open-wire pairs is due almost entirely to the external electric and magnetic fields of the disturbing circuit. If these fields were in some way annulled there would remain the possibility of resistance coupling between the pairs because of leakage from one circuit to the other by way of the crossarms and insulators, tree branches, etc. This leakage effect is minor in a well-maintained line. It enters as a factor in the design of open-wire transpositions only in so far as the attenuation of the circuits is affected which indirectly affects the crosstalk.

Figure 1 indicates cross-sections of two pairs of wires designated as 1–2 and 3–4. If pair 1–2 existed alone and if the two wires were similar, a voltage impressed at one end of the circuit would result in

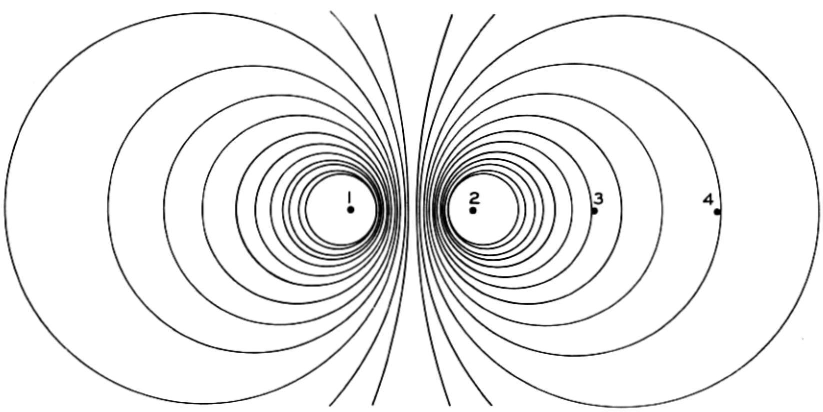

Fig. 1—Magnetic field produced by equal and opposite currents in wires 1 and 2.

equal and opposite currents at any point. These currents would produce a magnetic field as indicated on the figure. If circuit 3–4 parallels 1–2 a certain amount of this magnetic flux would thread between wires 3 and 4 and induce a voltage in circuit 3–4 which would result in a crosstalk current in this circuit. This induced voltage is, of course, due to the difference between the two magnetic fields set up by the opposite directional currents in wires 1 and 2. Since wires 1 and 2 are not very far apart, the resultant field is much weaker than if transmission over wire 1 with ground return were attempted. It is important, therefore, that the wires of a circuit be placed as close together as practicable and that these wires be similar in material and gauge in order to keep the currents practically equal and opposite.

Equal and opposite charges accompany the equal and opposite currents in wires 1 and 2. The equipotential lines of the resultant electric field set up by the two charges are also indicated by Fig. 1. This field will cause different potentials at the surfaces of wires 3 and 4 and this potential difference will cause a crosstalk current in circuit 3–4. As in the case of magnetic induction this current may be minimized by close spacing and electrical similarity between the two wires of a pair.

Calculations of crosstalk coupling must, in general, consider both the electric and magnetic components of the electromagnetic field of the disturbing circuit.

The exact computation of crosstalk coupling between communication circuits is very complex.[1] Approximate computations are sufficient for transposition design. In such computations, it is convenient to divide the total coupling into components of several general types. In calculations of coupling of these types it is assumed that the two wires of a circuit are similar in material and gauge. If there is any slight dissimilarity, such as extra resistance in one wire due to a poor joint, the effect on the crosstalk may be computed separately. The general types of crosstalk coupling are:

1. Transverse crosstalk coupling.
   1*a*. Direct.
   1*b*. Indirect.
2. Interaction crosstalk coupling.

A multi-wire pole line involves many circuits all mutually coupled. In explaining the above terms, it is convenient to start with the simple conception of but two paralleling coupled circuits; Fig. 2A

[1] The general mathematical theory is given in the Carson-Hoyt paper listed under "Bibliography."

indicates such a parallel. In calculating the crosstalk coupling between a terminal of circuit a and a terminal of circuit b, the parallel may be divided into a series of thin transverse slices. One such slice of thickness d is indicated on the figure. The coupling in each slice is calculated and, then, the total coupling between circuit terminals due to all the slices.

In Fig. 2A circuit a is considered to be the disturber and to be energized at the left-hand end. In the single slice indicated, a transmission current will be propagated along circuit a and will cause crosstalk currents in circuit b at both ends of the slice. In this slice, therefore, the left-hand end of circuit a may be considered to be coupled to the two ends of circuit b through the transmission paths n_{ab} and f_{ab}. The path n_{ab} is called the near-end crosstalk coupling and the path f_{ab} is called the far-end crosstalk coupling.

The presence of a tertiary circuit, such as c of Fig. 2B, changes both the near-end and the far-end coupling between a and b in the transverse slice. In addition to the direct couplings n_{ab} and f_{ab} there are indirect couplings n_{acb} and f_{acb} by way of circuit c.

The *transverse crosstalk coupling* between a terminal of a disturbing circuit and a terminal of a disturbed circuit is defined as the coupling between these points due to all the small couplings in all the thin transverse slices including indirect couplings in each slice by way of other circuits. (There are also indirect couplings involving more than one slice and these are not included in the transverse crosstalk coupling.)

In computations of transverse crosstalk coupling it is convenient to distinguish between the direct and indirect components. The direct component considers only the currents and charges in the disturbing circuit while the indirect component takes account of certain charges in tertiary circuits resulting from transmission over the disturbing circuit. The tertiary circuits may be circuits used for transmission purposes or any other circuits which can be made up of combinations of wires on the line or of these wires and ground. If there are only two pairs on the line as in Fig. 2A there are still tertiary circuits, namely, the "phantom" circuit consisting of pair a as one side of the circuit and pair b as the return and the "ghost" circuit consisting of all four wires with ground return. In a multi-wire line many of the tertiary circuits involve the wires of the disturbing circuit. If these tertiary circuits did not exist the currents at any point in the two wires of the disturbing circuit would be equal and opposite. The presence of the tertiary circuits makes these currents unequal and it is convenient to divide the actual currents into two components, i.e., equal and opposite or "balanced" currents in the two wires of the disturbing circuit and equal currents in phase in the two wires. The latter may be called "tertiary circuit" currents. The charges on the two wires of the disturbing circuit may be similarly divided into components.

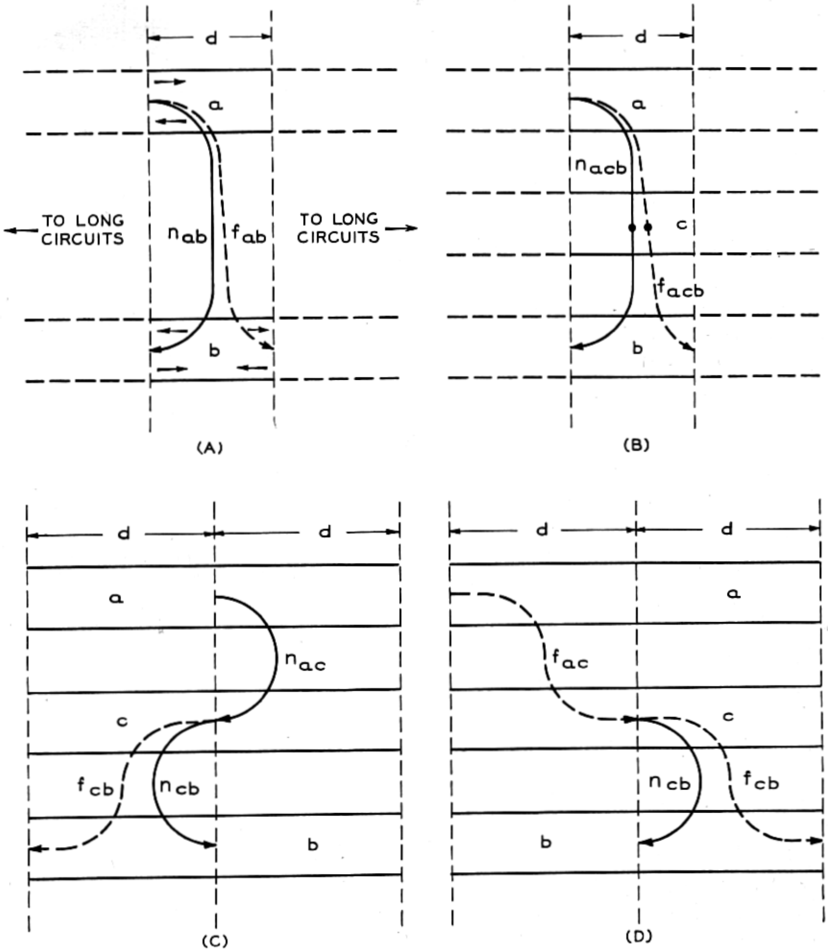

Fig. 2—Transverse and interaction crosstalk.

The *direct* component of the transverse crosstalk coupling is defined as that part which is due to balanced charges and currents in the disturbing circuit.[2] The *indirect* component is defined as that part

[2] "Direct" is here used in a different sense from that used in connection with the term "direct capacity unbalance" which was originated by Dr. G. A. Campbell and has been much used in discussions of cable crosstalk.

which is due to *charges* on tertiary circuits which arise within any thin transverse slice due to coupling with the disturbing circuit in that same slice. This coupling, in any slice, causes currents as well as charges in the tertiary circuit in that slice, but, as discussed in detail in Appendix A, the effect of these currents in producing crosstalk currents in the disturbed circuit is small compared with the effect of the charges. The currents and charges in the tertiary circuits in any thin slice due to the coupling with the disturbing circuit in that same slice may be but a small part of the total currents and charges in the slice. The total values are due to couplings of the tertiary circuits with the disturbing circuit in all the slices. When the total values are considered currents as well as charges in the tertiary circuit may be important in causing crosstalk currents in the disturbed circuit. To consider the total currents and charges in the tertiary circuits it is necessary to take account of both the interaction crosstalk coupling and the transverse crosstalk coupling between disturbing and disturbed circuits.

The nature of interaction crosstalk coupling is indicated by Figs. 2C and 2D which indicate two successive thin transverse slices of width d in a parallel between two circuits a and b and the typical tertiary circuit c. Assuming transmission from left to right on circuit a in Fig. 2C this circuit is coupled with c in the right-hand slice by the near-end crosstalk coupling indicated by n_{ac}. This coupling causes transmission of crosstalk current (and charge) into the left-hand part of circuit c which has both near-end and far-end crosstalk coupling to circuit b. Consideration of these two successive transverse slices, therefore, introduces the two compound couplings $n_{ac}n_{cb}$ and $n_{ac}f_{cb}$. There are two more of these compound couplings as indicated by Fig. 2D. There is a far-end crosstalk coupling between circuits a and c in the left-hand slice which combines with both near-end and far-end couplings in the right-hand slice. The compound types of crosstalk of Fig. 2C and 2D are called interaction crosstalk since the various slices interact on each other in producing indirect couplings. The *interaction crosstalk coupling* between a terminal of a disturbing circuit and a terminal of a disturbed circuit is defined as the coupling between these points due to the indirect couplings involving *all possible combinations* of different thin transverse slices.

The distinction between indirect transverse crosstalk and interaction crosstalk is that the former takes account of the effect of indirect crosstalk from disturbing to tertiary to disturbed circuit in a single thin transverse slice while the latter involves indirect crosstalk from primary circuit to tertiary circuit in one slice, transmission along the tertiary circuit into another slice and then crosstalk from tertiary circuit to disturbed circuit.

The notion that there is only transverse crosstalk within any one "thin slice" implies that the slice thickness corresponds to a distance along the line of only infinitesimal length. If this distance were finite it would correspond to a series of "thin slices" having interaction crosstalk between them. Practically, however, if the distance along the line corresponds to a line angle of five degrees or less, the interaction crosstalk in this length is small compared with the transverse crosstalk. A five degree line angle corresponds to a length of about .1 mile at 25 kilocycles, .05 mile at 50 kilocycles, etc. A transposed line is divided into short lengths or *segments* by the transposition poles and the line angle of these segments is ordinarily less than five degrees at the highest frequency for which the transposition system is suitable. Therefore, the crosstalk coupling between such transposed circuits may be computed on the basis of transverse crosstalk within any segment and interaction crosstalk between any two segments.

As shown by Fig. 2 the interaction effect involves the four compound couplings:

$$n_{ac}n_{cb}, \qquad n_{ac}f_{cb}, \qquad f_{ac}n_{cb}, \qquad f_{ac}f_{cb}.$$

The near-end crosstalk couplings n_{ac} and n_{cb} of Fig. 2 are usually much larger than the far-end couplings f_{ac} and f_{cb}. The reason for this, as discussed in Appendix A, is that the electric and magnetic fields of the disturbing circuit tend to aid each other in producing near-end crosstalk coupling such as n_{ac}, and to oppose each other in the case of far-end coupling such as f_{ac}. For this reason the compound coupling $n_{ac}n_{cb}$ is the most important and is usually the only compound coupling which requires consideration in transposition design. Since the path $n_{ac}n_{cb}$ results in a crosstalk current at the far end of the disturbed circuit, it is in connection with far-end crosstalk between long circuits that this matter of interaction crosstalk is important. Far-end rather than near-end crosstalk coupling is controlling in connection with open-wire carrier frequency systems for the reasons explained below.

Figure 3A indicates very schematically two one-way carrier frequency channels routed over two long paralleling open-wire pairs. The boxes at the end indicate the repeaters or terminal apparatus and the arrows on these boxes the direction of transmission of this apparatus. Transmission from the left on pair a results in near-end and far-end crosstalk into pair b, as indicated by the couplings n_{ab} and f_{ab}. The near-end crosstalk current cannot pass to the input of the terminal apparatus since the latter is a one-way device. In practice, to obtain two-way circuits each of these one-way channels is associated with another one-way channel transmitting in the opposite direction over the same pair of wires. These return channels utilize a different band of carrier frequencies and the near-end crosstalk current is largely excluded from this frequency band by selective filters. The far-end crosstalk is, therefore, the sole consideration with such a carrier system. Use is not made of the same carrier frequencies in both directions on a toll line largely because of difficulties in controlling the near-end crosstalk.

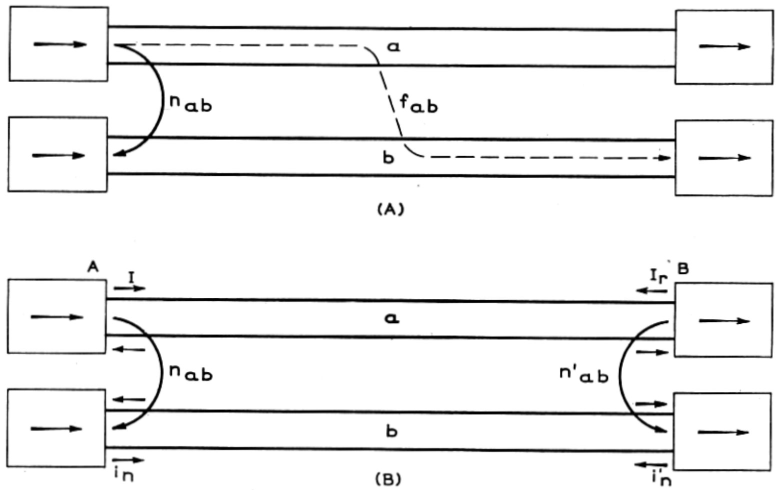

Fig. 3—Crosstalk between two one-way carrier frequency channels.

In connection with the arrangement of Fig. 3A, there is a type of crosstalk of considerable practical importance known as "*reflection crosstalk*." The theory of this is indicated by Fig. 3B which shows the same two one-way carrier channels. Transmission from left to right on circuit a is assumed. When the transmission current I arrives at point B, a certain portion of it will be reflected if there is any deviation of the input impedance of the terminal apparatus from the characteristic impedance of circuit a. This reflected current I_r causes a near-end crosstalk current i_n' at point B in the disturbed circuit. Similarly, a part of the near-end crosstalk current i_n at point A in the disturbed circuit may be reflected and transmitted to point B. Therefore, two additional crosstalk currents may result from these two reflections and such currents can enter the terminal apparatus at B and pass through to the output of this apparatus.

For like circuits, like impedance mismatches and like near-end crosstalk couplings at the two ends of the line, these two additional far-end crosstalk currents are of equal importance. Similar reflection effects will occur at any intermediate points in the lines having impedance irregularities. Since the far-end crosstalk coupling can be much more readily reduced by transpositions than the near-end crosstalk coupling this reflection crosstalk effect is important in practice. It is, therefore, necessary to carefully design the terminal and intermediate apparatus and cables to minimize impedance mismatches as far as practicable.

In calculation of crosstalk coupling it is ordinarily assumed that the two wires of a circuit are electrically similar or "balanced" (except as regards crosstalk from other wires). This is substantially true in practice except for accidental deviations, such as resistance differences due to poor joints and leakage differences due to cracked insulators, foliage, etc. Resistance differences may be of considerable practical importance and are said to cause *resistance unbalance crosstalk.* The following discussion indicates the general nature of this effect.

As discussed in connection with Fig. 1, the external field of the disturbing circuit is minimized by the opposing effects of substantially equal and opposite currents or charges in the two wires of the circuit. The two wires may be considered as two separate circuits, each having its return in the ground. At any point in the line these two wires would normally have practically equal and opposite voltages with respect to ground. These voltages would normally cause almost equal and opposite currents in the two wires. If the resistance of one wire is increased due to a bad joint, the current in that wire is reduced and the currents in the two wires are no longer equal and opposite. The external field of the two wires and the resulting voltage induced in the disturbed circuit are, therefore, altered. If this voltage had previously been practically cancelled out by means of transpositions, the alteration in the field would increase the crosstalk current at the terminal of the disturbed circuit.

A resistance unbalance in the disturbed circuit will have a similar effect as indicated by Fig. 4A. This figure shows a short length d of two long paralleling circuits. Equal and opposite transmission currents in the disturbing circuit 1–2 are indicated by I. Equal crosstalk currents in the two wires of the disturbed circuit 3–4 at one end of the short length are indicated by i. It is assumed that these crosstalk currents have been made substantially equal by transpositions in other parts of the line. Since the currents in wires 3 and 4 are equal and in the same direction, there will be no current in a receiver connected at the terminal of the line between these wires. If, however, one wire has a bad joint, the two crosstalk currents become unequal and there will be a current in such a receiver.

Resistance unbalance crosstalk is of particular importance if two pairs are used to create a phantom circuit in order to obtain three transmission circuits from the four wires. The distribution of the phantom transmission current I_p in a short length of the two pairs is indicated by Fig. 4B. Ideally, half the phantom current flows in each of the four wires. The two currents in wires 1 and 2 are then equal and in the same direction and there will be no current in terminal apparatus connected between wires 1 and 2. In other words, transmission over the phantom circuit results in no crosstalk in the side circuit 1–2. The same may be said of side circuit 3–4. A bad joint in any wire, such as 3, makes the two currents in wires 3 and 4 unequal and results in a current in the side circuit 3–4.

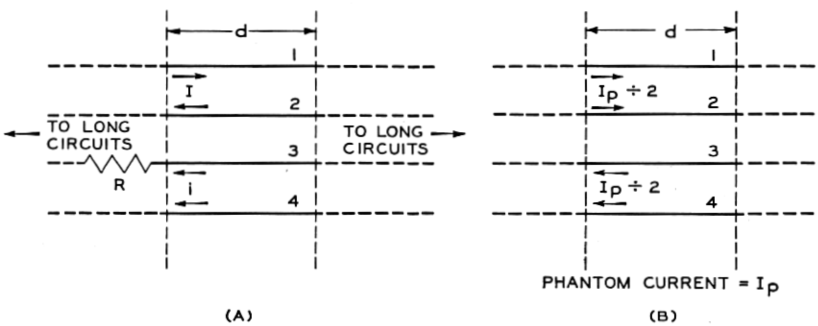

Fig. 4—Effect of resistance unbalance on crosstalk.

The phantom-to-side crosstalk effect of resistance unbalance is much more severe than the effect on crosstalk between two side circuits or two non-phantomed circuits. The reason for this is evident from Figs. 4B and 4A. In Fig. 4B the entire transmission current of the disturbing phantom circuit normally flows in the two wires of the disturbed side circuit and if a resistance unbalance causes a small percentage difference in the currents in these two wires objectionable crosstalk results. In Fig. 4A only crosstalk currents flow in wires 3 and 4 and a much larger percentage difference between these small currents can be tolerated.

In designing and operating phantom circuits, it is necessary to exercise great care to minimize any dissimilarity between the two wires of a side circuit, in order to avoid crosstalk from a phantom to its side circuit or vice versa. Otherwise, the problem of crosstalk between a phantom circuit and some other circuit is generally similar to the problem of crosstalk between two pairs. In other words, the discussion of transverse, interaction and reflection crosstalk is applicable.

Measures of Crosstalk Coupling

In designing transposition systems, the usual measure of the coupling effect between two open-wire circuits is the ratio of current at the output terminal of the disturbed circuit to current at the input terminal of the disturbing circuit. For circuits of different characteristic impedances this current ratio must be corrected for the difference in impedance. The corrected current ratio is the square root of the corresponding power ratio.

The current ratio is ordinarily very small and for convenience is multiplied by 1,000,000 and called the crosstalk coupling or, in brief, the crosstalk. This usage will be followed from this point in this paper. For example, crosstalk of 1000 units means a current ratio of .001. Crosstalk may also be expressed as the transmission loss in db corresponding to the current ratio. A ratio of .001 means a transmission loss of 60 db corresponding to 1000 crosstalk units.

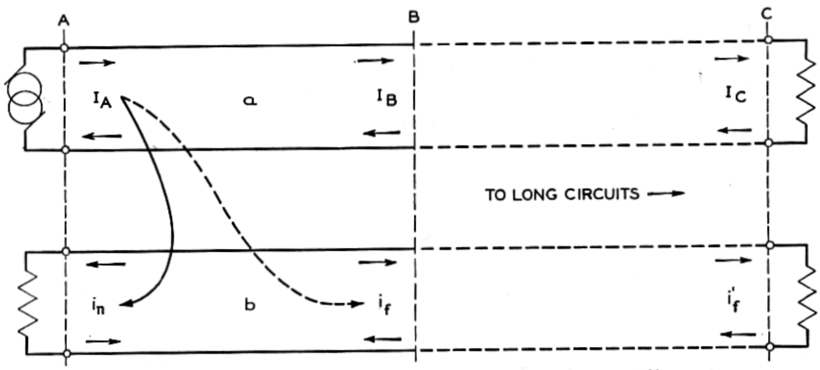

Fig. 5—Schematic of near-end and far-end crosstalk.

Figure 5 indicates two paralleling communication circuits *a* and *b* with an e.m.f. impressed at one end of circuit *a*. The crosstalk currents i_n and i_f in circuit *b* are due to the crosstalk coupling in length AB. The *near-end crosstalk* in the length AB is the ratio $10^6 i_n/I_A$, while the *far-end crosstalk* is $10^6 i_f/I_A$. The ratio $10^6 i_f/I_B$ has been called the "output-to-output" or "measured" crosstalk. This ratio is a convenient measure of far-end crosstalk between parts of similar circuits because it is related in a simple way to the far-end crosstalk between the terminals of the complete circuits. The following discussion explains this relation.

Both of the currents I_B and i_f will be propagated to point C. They will be attenuated or amplified alike if the circuits are similar and their ratio will be unchanged. The output-to-output crosstalk at C due to the length AB will, therefore, be the same as that determined for point B. In other words $10^6 i_f'/I_C$ will equal $10^6 i_f'/I_B$. The far-end crosstalk between the terminals A and C, due to length AB, will be $10^6 i_f'/I_A$. This differs from the output-to-output crosstalk at C in that the reference current is I_A instead of I_C. The part of the far-end crosstalk between A and C due to AB is, therefore, obtained from the output-to-output crosstalk at B by simply multiplying by the attenuation ratio I_C/I_A. If the output-to-output crosstalk is expressed as a loss in decibels, the far-end crosstalk is obtained by adding the net loss of the complete circuit between A and C.

Effects of Transpositions

The effects of transpositions on both the transmission currents and the crosstalk currents will now be discussed in a general way. The general method of computing the crosstalk between circuits without constructional irregularities and transposed in any manner will also be outlined.

General Principles

If there is only one circuit on a pole line, and this is balanced and free from irregularities, the communication currents will be propagated along this circuit according to the simple exponential law. If a current is propagated from the start of the circuit to some other point at a distance L, the magnitude of the current will be reduced by the attenuation factor $\epsilon^{-\alpha L}$ and the phase of the current will be retarded by the angle βL where α is the attenuation constant and β is the phase change constant.

If there are a number of circuits on a pole line this simple law of propagation may be altered due to crosstalk into surrounding circuits. This is illustrated by the curves, Fig. 6, which indicate the relation between observed output-to-input current ratio and frequency for two different circuits, each about 300 miles long and having 165-mil copper wires. The number of decibels corresponding to the current ratio is plotted rather than the ratio itself. For the simple law of propagation such curves would show the number of decibels increasing smoothly with frequency due to increasing losses in the line wires and insulators. The upper curve is for a circuit too infrequently transposed for the frequency range covered and the current ratio is abnormally small at particular frequencies. The corresponding number of decibels is abnormally large. The lower curve is for a circuit much more frequently transposed and its current ratios practically follow the simple propagation law mentioned above over the frequency range shown.

Even though a circuit is very frequently transposed, its propagation constant is slightly affected by the presence of other circuits on the line. This may be explained by consideration of Figs. 2B and 7. As previously explained, Fig. 2B indicates the indirect transverse crosstalk by way of a tertiary circuit in one thin transverse slice of a parallel

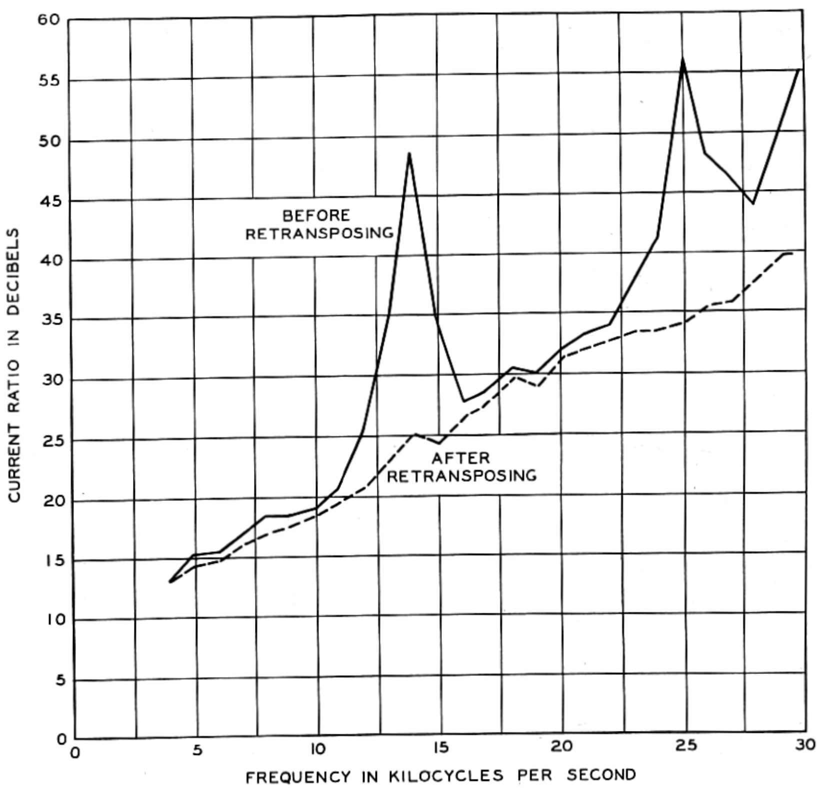

Fig. 6—Effect of transpositions on attenuation of an open-wire pair.

between two long circuits *a* and *b*. The circuit *c* has currents and charges due to crosstalk from the disturbing circuit *a*. These currents and charges not only alter the crosstalk currents in circuit *b* but also react to change the transmission current in circuit *a*. Since circuits *a* and *c* are loosely coupled, this reaction effect could usually be estimated with sufficient accuracy by calculating the crosstalk from *a* to *c* and back again and neglecting the further reactions of the change in the current in *a* on the current in *c*, etc.

Figure 7 shows the crosstalk paths from *a* to *c* and back again. In this figure, circuit *a* is indicated as two separate circuits for comparison with Fig. 2B. It is assumed that circuit *a* in Fig. 7 is energized at point *A*, the currents I_A and I_B being the currents which would exist at the input and output of the short length *d* if there were no tertiary circuits. The near-end crosstalk path indicated by *n* will cause a small crosstalk current i_n at point *A* in circuit *a*. There will be a crosstalk path similar to *n* in each thin slice of the parallel between *a* and *c*. Each of these paths will transmit a small crosstalk current to point *A* in circuit *a*. The sum of all these crosstalk currents will *increase* the input current I_A and, therefore, the impedance of circuit

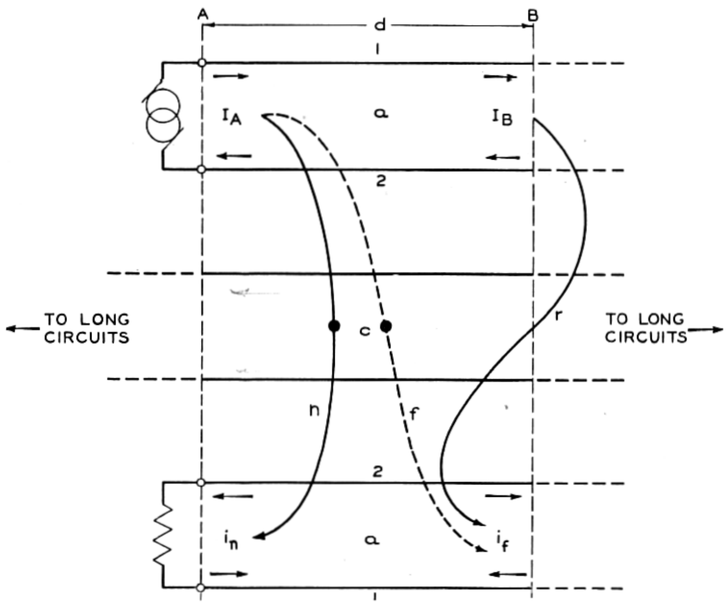

Fig. 7—Effect of circuit c on propagation in circuit a.

a is lowered. Thin slices remote from the sending end will contribute little to this effect, since the crosstalk currents from such slices will be attenuated to negligible proportions. A long circuit on a multi-wire line will, therefore, have a definite sending-end impedance slightly lower than that for one circuit alone on the line.

Figure 7 also indicates a far-end crosstalk path *f* which produces a crosstalk current i_f at point *B* in circuit *a*. This *reduces* the transmission current I_B at this point and, therefore, increases the attenuation constant of the circuit. For calculations of both the circuit impedance and attenuation, the effect of surrounding circuits is taken care of in practice by using a capacity per unit length slightly higher than the value which would exist with only one circuit on the line. The proper capacity to use is determined in practice by measurements on a short length of a multi-wire line.

The effect on the propagation constant of the transverse crosstalk paths indicated by n and f of Fig. 7 cannot be suppressed by transpositions. As explained later, if the two circuits marked a were actually different circuits, the effect could be largely suppressed by transposing one circuit at certain points and leaving the other circuit untransposed at these points. Since the disturbing and disturbed circuits indicated by Fig. 7 are actually the same circuit, they must be transposed at the same points and, therefore, the transverse crosstalk effect cannot be suppressed by frequent transpositions.

Figure 7 also shows a crosstalk path marked r. This is one of the possible interaction crosstalk paths. The effect of such paths on the impedance and attenuation of the circuit may be largely suppressed by suitable transpositions. The difference between the two curves of Fig. 6 is due to lack of this suppression in the case of the upper curve.

Such an extreme effect of crosstalk reacting back into the primary or initiating transmission circuit and thus affecting direct transmission is seldom important in practical transposition design. A marked reaction on the primary circuit would necessitate such large crosstalk currents in neighboring communication circuits as to make them unfit for communication service at the frequency transmitted over the primary circuit. Therefore, it is only when the neighboring circuits are not to be used at this frequency that transposition design to control simply the direct transmission becomes of practical importance. When many circuits on a line are used for carrier operation, the crosstalk currents must be made so weak (by transpositions, physical separation of circuits, etc.) that their reactions back into the primary circuits are very small.

The effect of transpositions on crosstalk from one circuit into another *different* circuit will now be considered. The discussion of the control of this effect is the main object of this paper.

Figure 8A shows a short segment of a parallel between two long circuits and a near-end crosstalk coupling marked n. The segment could be divided into a series of thin slices and theoretically there would be interaction crosstalk between different slices. The segment length is, however, assumed to be short enough to neglect interaction crosstalk. The coupling n is, therefore, due either to direct or indirect transverse crosstalk in the short segment or to both of these types of crosstalk. If circuit a is energized from the left, a near-end crosstalk current i_n results at point A in circuit b.

If two successive short segments are considered, as indicated by Fig. 8B, there will be a near-end crosstalk coupling n in each segment and each of these couplings will result in a crosstalk current at point A

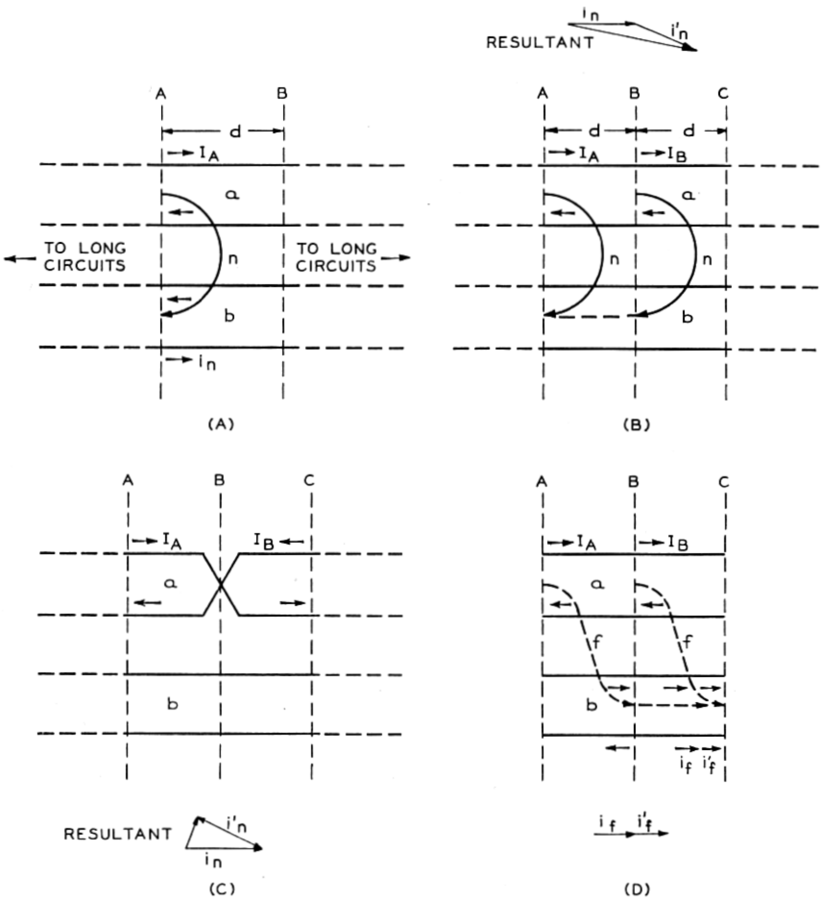

Fig. 8—Effect of transpositions on transverse crosstalk.

of circuit b. This is indicated by the vector diagram over the figure, where i_n indicates the crosstalk current due to the segment AB and i_n' indicates the crosstalk current at A due to BC. The latter current is slightly smaller and slightly retarded in phase with respect to i_n because in order for i_n' to appear at point A, the transmission current I_A must be propagated a distance d and the resulting crosstalk current at B must also be propagated a distance d in order to reach A. As indicated by this vector diagram the total crosstalk current due to the two short segments is a little less than the arithmetic sum of the individual crosstalk currents.

Figure 8C is like Fig. 8B except that a transposition is inserted in the middle of circuit a at point B. This reverses the phase of the transmission current at the right of B and also reverses any crosstalk current due to current in circuit a between B and C. As a result the crosstalk current i_n' of Fig. 8B is reversed and the resultant of the two crosstalk currents is very much reduced as indicated by the vector diagram of Fig. 8C. The angle between i_n and i_n' is proportional to the length $2d$ which equals AC. The tendency for the two currents to cancel may, therefore, be increased by reducing the length AC which, in a long line, would mean increasing the number of transpositions.

Figure 8D is like Fig. 8B except that the far-end transverse crosstalk coupling f in each of the two short segments is considered. The coupling in the left-hand segment results in a crosstalk current at point B of circuit b, which is propagated to point C as indicated by i_f. The far-end crosstalk coupling in the right-hand segment produces a crosstalk current i_f' at point C. Since the total propagation distance is from A to C for both of these crosstalk currents, they must be equal in magnitude and in phase if circuits a and b are similar. This is indicated by the vector diagram of Fig. 8D. A transposition at point B in either circuit would reverse one of these crosstalk currents and, therefore, the resultant crosstalk current would be nil.

From consideration of Figs. 8C and 8D, it may be seen that if both circuits were transposed at point B, the sum of the crosstalk currents for the two segments would be the same as if neither circuit were transposed. Transposing one circuit reverses the phase of one of the component crosstalk currents, but if the second circuit is also transposed the original phase relations between the two currents are restored.

The foregoing discussion applies only to transverse crosstalk as discussed in connection with Fig. 2. When interaction crosstalk must be considered, a different principle is involved.

In connection with Fig. 8D, it was shown that the transverse far-end crosstalk between similar circuits could be readily annulled by transposing one of the circuits at the center of their paralleling length. Far-end crosstalk of the interaction type is not so readily annulled. The effect of transpositions on this type of crosstalk is indicated by Fig. 9.

This figure shows four short segments in a parallel between two circuits a and b, there being an interposed tertiary circuit c. Interaction crosstalk involving two near-end crosstalk couplings is considered since this is usually the controlling type. There is an interaction crosstalk path designated r between the first two segments as indicated by Fig. 9A. There is a similar path between the third and fourth segments. Each of these paths would produce a far-end crosstalk current in circuit b at point E. For similar circuits these currents would be equal in magnitude and would add directly. The two currents can be made to cancel by transposing one of the circuits at C, the midpoint of the parallel. Such a transposition also cancels the transverse far-end crosstalk in length AC against that in length CE. There remains, however, the interaction crosstalk between length CE and length AC.

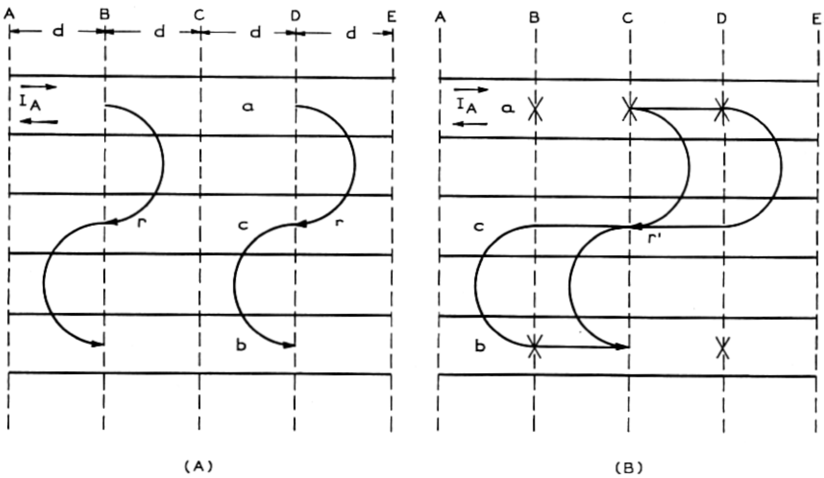

Fig. 9—Effect of transpositions on interaction crosstalk.

Figure 9B shows a transposition at C in circuit a and also other transpositions whose purpose is to minimize the interaction crosstalk between length CE and length AC. This crosstalk coupling, designated by r', is a compound effect, depending on the near-end crosstalk between circuit a and circuit c in length CE and the near-end crosstalk between c and b in length AC. The near-end crosstalk coupling between a and c in length CE can be greatly reduced by a transposition in circuit a at point D, while the crosstalk coupling between c and b in length AC can likewise be reduced by a transposition at point B in circuit b. The latter two transpositions would not, however, minimize the interaction crosstalk between CE and AC with circuit b as the disturbing circuit and it is necessary, therefore, to transpose both circuits at points B and D. The addition of these four transpositions does not affect the cancellation of far-end crosstalk in length AC against that in length CE by means of the transposition at C. After the four transpositions are added, length AC is still similar to length CE and the far-end crosstalk currents at E, due to these two lengths, are equal. Therefore, they will cancel when one of them is reversed in phase by the transposition at C.

It may be concluded that, while transposing both circuits at the same points has no effect on transverse crosstalk, it has a large effect on the interaction crosstalk. An experimental illustration is given in Fig. 10. This figure shows frequency plotted against output-to-output

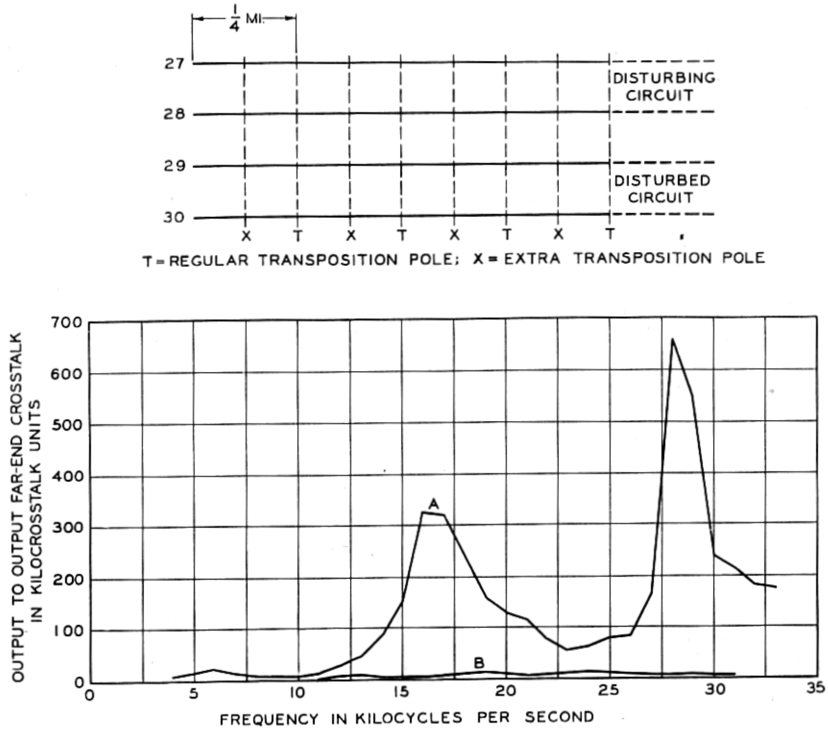

Fig. 10—Effect on far-end crosstalk of extra transpositions in both circuits.

far-end crosstalk between the two side circuits of a phantom group on a 140-mile length of line. The curve marked A is for the two circuits transposed for voice-frequency operation. Curve B is for the two circuits transposed in the same manner except that four transpositions per mile were added to both circuits at the same points which are indicated by x on Fig. 10. The large effect of these transpositions shows the practical importance of the interaction type of far-end crosstalk.

In connection with Fig. 9B, there arises the question of how far apart the transpositions can be placed without serious crosstalk, in other words, how long is it permissible to make the segment d. If this length is increased the transpositions at B and D become less effective in suppressing the near-end crosstalk between a and c in length CE and between c and b in length AC. The degree to which the interaction crosstalk path r' must be suppressed is, therefore, important in determining the maximum permissible length of d. If d is increased the transposition at C becomes less effective in controlling the near-end crosstalk between a and b and, therefore, the length d also depends on the permissible near-end crosstalk.

It may be noted that transpositions at B and D in but one of the circuits a or b will help to suppress r', but the suppression is less effective than if both circuits are transposed at these points. If a is transposed at B and D the near-end crosstalk between a and c in length CE is reduced but the near-end crosstalk between c and b in length AC is not reduced. The product of these two near-end crosstalk values is greater, therefore, than if they had both been reduced by transposing both circuits at B and D.

Crosstalk Coefficients

The crosstalk between any two long open-wire circuits may be calculated by dividing the parallel into a succession of thin transverse slices and summing up the crosstalk for all these slices. To calculate the crosstalk in any slice it is necessary to know certain "crosstalk coefficients." The discussion below defines these coefficients and describes briefly how they are measured or computed.

Figures 2A and 2B indicate both near-end and far-end crosstalk coupling of both the direct and indirect transverse types in a thin transverse slice. Any of these couplings may be expressed in crosstalk units and the value of the coupling in a short length divided by the length in miles is called the crosstalk per mile. Since, as shown in the previous section, the crosstalk may not increase directly as length, strictly speaking, the crosstalk per mile is the limit of the ratio of coupling to length as the length approaches zero. The crosstalk per mile includes both the direct and indirect types of transverse crosstalk coupling. In the frequency range of interest (i.e., above a few hundred cycles for near-end crosstalk and above a few thousand cycles for far-end crosstalk) this total transverse coupling varies about directly with the frequency and the crosstalk coefficient commonly used is the *crosstalk per mile per kilocycle.*

If many wires are involved, it is impracticable to determine these coefficients with good accuracy by computation and they are, therefore, derived from measurements. Examples of near-end and far-end coefficients, plotted against frequency, are shown in Fig. 11. The coefficients are for pairs designated 1–2 and 3–4 on the pole head diagram shown on the figure. These coefficients were derived from measurements of the near-end and far-end crosstalk over a range of frequencies. The length of line was about .2 mile and, for the range of frequencies covered, this length is sufficiently short so that interaction crosstalk is negligible and the transverse crosstalk is directly proportional to the length. The coefficients plotted are, therefore, nearly equal to the measured values of crosstalk divided by the length and by the frequency. (A small correction was made at the higher frequencies to allow for deviation of near-end crosstalk from simple proportionality to length and the curves were "smoothed" through the actual points calculated from the measurements.)

In order to obtain the crosstalk coefficients applicable to a short part of a long line, all the wires on the line were terminated in such a manner as to roughly simulate their extension for long distances in both directions, but without crosstalk coupling between the test pairs in such extensions. This is done by terminating each pair at each end with a resistance approximating its characteristic impedance and connecting the midpoint of each resistance to ground through a second resistance. These latter resistances terminate any phantom of two pairs as indicated on Fig. 11 for pairs 1–2 and 3–4. Any circuit with ground return is also terminated by these resistances.

Both of the test pairs are transposed at the midpoint of the line during the measurement. This minimizes the currents reaching the ends of the tertiary circuits and makes even the above approximate termination of the tertiary circuits of little importance.

Figure 11 shows near-end and far-end crosstalk coefficients for three conditions, A, B, and C. The two curves marked A show the measured values with all wires terminated and the test pairs transposed as described above.

For curves B, only the transposed test pairs were terminated as described above and the other wires were opened at the middle, at the quarter points and at both ends. Since no section of any of these wires connected points of substantially different potential in the field of the disturbing circuit there were practically no currents or charges in these wires and the crosstalk coefficients for the two test pairs were practically the same as if the other wires had been removed from the line. It will be seen that the crosstalk coefficients for curves *B* are

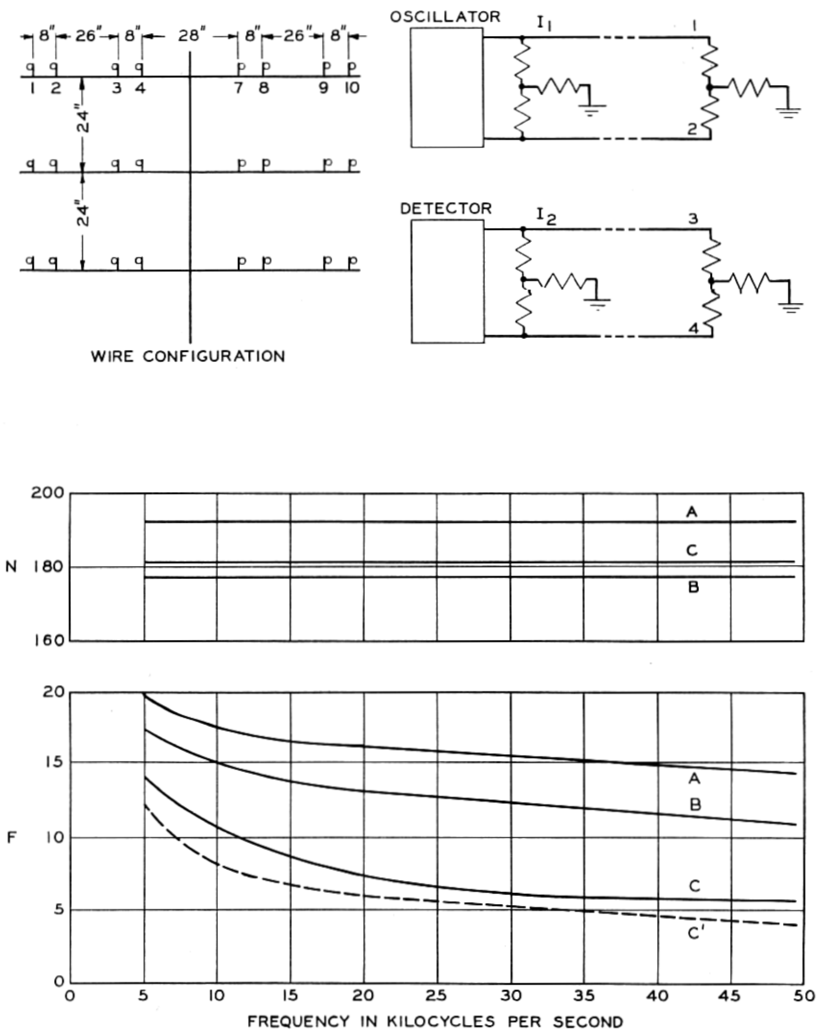

Fig. 11—Near-end and far-end crosstalk coefficients between pairs 1–2 and 3–4.

less than those for curves *A*. The coefficients of curves *B* involve tertiary circuits, however, since there could be crosstalk currents in the phantom of the two test pairs and also in the ghost circuit involving wires 1 to 4 with ground return.

Curves C show the coefficients with the test pairs *without* transpositions and terminated at both ends as accurately as practicable, but without the midpoints of these terminations connected to ground to terminate the phantom and ghost circuits. These tertiary circuits were, with this arrangement, prevented from connecting points of substantially different potential and the coefficients of curves C, therefore, approach the direct crosstalk coefficients. It is extremely difficult to experimentally determine the direct far-end coefficient. It may be computed, however, and the computed value which assumes perfect terminations and the effect of the phantom completely removed is shown by curve C'.

It may be noted that the near-end crosstalk coefficients are about independent of frequency. This is ordinarily true above a few hundred cycles. The total far-end coefficient (curve A) is about independent of frequency in the important carrier frequency range. The direct far-end coefficient of curve C' decreases considerably with frequency for reasons discussed in Appendix A. Since transpositions are ordinarily designed for the condition of a number of wires on a line, the total crosstalk coefficient is the one usually used in practice.

Curves C of Fig. 11 also indicate that the direct near-end coefficient is much larger than the direct far-end coefficient. This is usually true and, as discussed in detail in Appendix A, the explanation is that the crosstalk currents caused by the electric and magnetic fields add almost directly in the case of direct near-end crosstalk but tend to cancel in the case of direct far-end crosstalk. As discussed in the appendix, the indirect (vector difference of curves A and C) crosstalk in a very short length is due almost entirely to the electric field of the tertiary circuits and is the same for both near-end and far-end crosstalk. In Fig. 11, the total near-end coefficient (curve A) is increased by the indirect crosstalk since curve C is lower than curve A. The reverse is usually true, however. In the case of far-end crosstalk the total coefficient is usually increased by the indirect crosstalk.

Crosstalk coefficients are vector quantities and may be measured in magnitude and phase. If it is desired to compute the crosstalk between two long pairs of wires which do not change their pin positions, it is only necessary to know the magnitude of the crosstalk coefficient, since the problem is to determine the ratio of the crosstalk for many elementary lengths to the crosstalk for one such length. However, if it is desired to know the crosstalk between long circuits which do change their pin positions, several crosstalk coefficients must be known, one for each combination of pin positions. In order to determine the total crosstalk for several segments of a line involving different pin positions, it is necessary to know both the phase and magnitude of the crosstalk coefficients. For practical purposes, however, the coefficients may, in most cases, be regarded as algebraic quantities having sign but not angle.

The direct component of the total crosstalk coefficient may be readily computed as discussed in Appendix A. If more than a very few wires are involved, an exact calculation of the indirect component is impracticable but a fair approximation may be obtained by the method discussed in Appendix A. This method is used when a wire configuration is under consideration but is not available for measurement.

As pointed out in 1907 by Dr. G. A. Campbell, an accurate calculation of the total crosstalk coefficient would involve determination of the "direct capacitances" between wires of the test pairs. Since these capacitances are functions of the distances between all combinations of wires on the lead and between wires and ground, their calculation is usually impracticable. In the past, the crosstalk coefficients were computed by a method proposed by Dr. Campbell which involved measurement of the direct capacitances.[3]

The part of the coefficient due to the electric field was computed from the "direct capacitance unbalance." The part due to the magnetic field was computed as discussed in Appendix A. When loaded open-wire circuits were in vogue it was necessary to be able to separate the electric and magnetic components of the coefficients. After loading was abandoned this separation was unnecessary and it was found more convenient to measure the total coefficients than to measure the direct capacitances or differences between pairs of these capacitances.

As previously discussed, in designing transpositions it is necessary to compute the interaction type of crosstalk indicated by Fig. 2C, and it is, therefore, necessary to have some coupling factor for use in this computation. Such a coupling factor could, theoretically, be determined as indicated schematically by Fig. 12. The interaction crosstalk between two short lengths of line would be measured by transmitting on one pair and receiving on the other pair at the junction of the two short lengths as indicated by the figure.

If there were but a single tertiary circuit such as c of the figure, the crosstalk measured would be that due to the compound crosstalk path $n_{ac}n_{cb}$. In this product, n_{ac} is the near-end crosstalk between a and c in the right-hand short length d and n_{cb} is the near-end crosstalk between c and b in the left-hand short length. Since n_{ac} and n_{cb} when

[3] See papers by Dr. Campbell and Dr. Osborne listed under "Bibliography."

expressed in crosstalk units are current ratios times a million, their product $n_{ac}n_{cb}$ is a current ratio times a million squared. The crosstalk measured would be this current ratio times a million or $n_{ac}n_{cb}10^{-6}$.

For small values of d, n_{ac} and n_{cb} vary directly as the frequency and as the length d. Therefore:

$$n_{ac} = N_{ac}Kd,$$

$$n_{cb} = N_{cb}Kd,$$

$$n_{ac}n_{cb}10^{-6} = N_{ac}N_{cb}K^2d^210^{-6},$$

where N_{ac} and N_{cb} are the near-end crosstalk coefficients, K is the frequency in kilocycles and d is expressed in miles. The measured

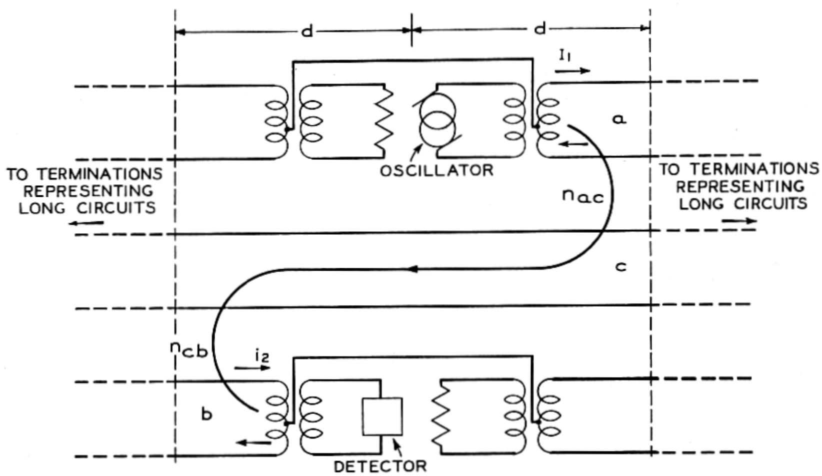

Fig. 12—Theoretical method of measuring interaction crosstalk coefficient.

crosstalk $n_{ac}n_{cb}10^{-6}$ divided by K^2d^2 gives the quantity $N_{ac}N_{cb}10^{-6}$ which may be designated as I_{ab} and called *the interaction crosstalk coefficient.* Values of I_{ab} determined from crosstalk measurements on multi-wire lines would include the effect of numerous tertiary circuits instead of that of a single tertiary circuit as indicated by Fig. 12.

While the interaction crosstalk coefficient I_{ab} could theoretically be measured as outlined above, it is simpler to deduce an approximate value from the measured value of the far-end crosstalk coefficient F_{ab}. The indirect component of F_{ab} is due to the tertiary circuits and must, therefore, be related to I_{ab} which is also due to these circuits. As discussed in detail in Appendix A:

$$I_{ab} = -\frac{2\gamma_c F_{ab}}{K} \text{ approximately.}$$

In this expression K is the frequency in kilocycles and $\gamma_c = \alpha_c + j\beta_c$ is the propagation constant of the tertiary circuit c. On a multi-wire line there would be numerous tertiary circuits with various values of γ. With practicable wire sizes the attenuation constants indicated by α are small compared with the phase change constants indicated by β. Measurements of crosstalk indicate that the values of β are all in the neighborhood of the value given by the expression $\pi K/90$. This corresponds to a speed of propagation of 180,000 miles per second which is about the average for the present carrier frequency range. Neglecting the attenuation constants:

$$\gamma_c = j\beta = j\frac{\pi K}{90},$$

$$I_{ab} = -j\frac{2\pi F_{ab}}{90}.$$

This relation is much used in transposition design. As noted above, the indirect component of F_{ab} should, strictly speaking, be used to obtain I_{ab}. In most cases, however, the total value of F_{ab} may be used since this total is determined largely by the indirect component.

Type Unbalance

A conception important in transposition design is that of "type unbalance." This conception will now be explained and the general method of computation will be discussed.

As we have seen, any two open-wire circuits tend to crosstalk into each other due to coupling between them. By transposing the circuits, the coupling in any short length of line is nearly balanced in another short length by a second coupling of about the same size but about opposite in phase. This balancing is never perfect and there is always a residual unbalanced coupling due to (1) attenuation and change in phase of the disturbing transmission current and resulting crosstalk currents as they are propagated along the circuits and (2) irregularities in the spacing of the transpositions and irregularities in the spacings between the various wires. The term "type unbalance" has been chosen to indicate the residual unbalance caused by propagation effects. It is expressed as an "equivalent untransposed length," that is, the type unbalance times the crosstalk per mile gives the residual crosstalk due to propagation effects assuming no constructional irregularities.

The method of computing the type unbalance for near-end crosstalk will now be discussed. The part of the near-end crosstalk due to interaction between all the different thin slices of line may be ignored since, as discussed in connection with Figs. 2C and 2D, the interaction crosstalk involves the product of a near-end crosstalk path and a far-end crosstalk path. This product is small since the coupling through the far-end path is inherently small. Therefore, the interaction crosstalk coefficient is much smaller for near-end crosstalk than for far-end crosstalk, while for the transverse crosstalk coefficients the reverse is true.

As was indicated by the discussion of Fig. 8B, the transverse near-end crosstalk between two long circuits may be computed by dividing the parallel into short segments, each having the same transverse crosstalk coupling. The coupling between circuit terminals for any segment will be different from that at the segment terminals due to propagation effects as explained in connection with Fig. 8B. Therefore, the coupling at the circuit terminal for each segment must be determined and, finally, the sum of the coupling values for all the segments.

The simplest case is that of two non-transposed circuits. The problem is indicated by Fig. 13 which is like Fig. 8B except that more segments are shown.

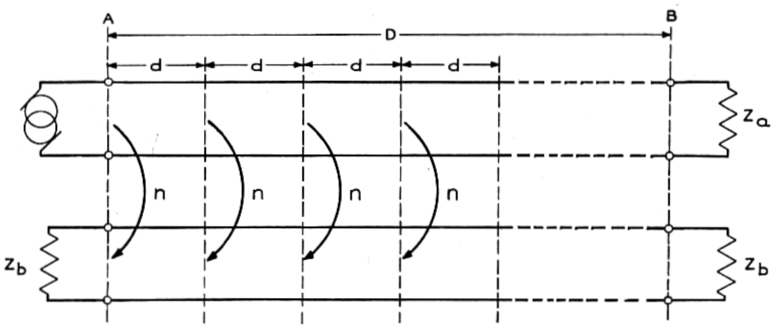

Fig. 13—Method of computing near-end crosstalk between untransposed circuits in length D.

The near-end crosstalk coupling n at point A due to the first segment is NKd, where N is the crosstalk coefficient and K is the frequency in kilocycles. The crosstalk current from the second segment relative to that from the first segment is attenuated by the factor $\epsilon^{-(\alpha_1+\alpha_2)d}$, and also retarded in phase by the angle $\epsilon^{-j(\beta_1+\beta_2)d}$. In other words, the crosstalk current from the second segment is equal to the crosstalk current from the first segment times the factor $\epsilon^{-(\gamma_1+\gamma_2)d}$, where γ_1 and γ_2 are the propagation constants for the two circuits and γ equals $\alpha + j\beta$. Letting γ be the average propagation constant, the coupling at point A for the second segment is equal to that for the first segment times $\epsilon^{-2\gamma d}$ or $NKd\epsilon^{-2\gamma d}$. The coupling at point A for the third segment is $NKd\epsilon^{-4\gamma d}$. The sum of the crosstalk couplings at point A at all the segments is, therefore:

$$NKd(1 + \epsilon^{-2\gamma d} + \epsilon^{-4\gamma d} + \epsilon^{-6\gamma d} + \text{etc.}).$$

This expression may be summed up for the number of segments corresponding to the total length D. It is simpler, however, to let d be an infinitesimal length and to integrate over the length D, i.e., from point A to point B of Fig. 13. This gives for the total near-end crosstalk for non-transposed circuits:

$$NK\frac{1 - \epsilon^{-2\gamma D}}{2\gamma}.$$

In the special case when D is only the usual short segment between transposition poles, the above expression is practically equal to NKD.

The near-end crosstalk between circuits having transposition poles spaced a considerable distance D apart may now be computed. Figure 14 shows a length $2D$ in a parallel between two long circuits, there being a transposition in one circuit at the center of $2D$. The near-end crosstalk for the length AB is given by the above expression. The near-end crosstalk at point A for the length BC will be the same expression multiplied by the propagation factor $\epsilon^{-2\gamma D}$ and reversed in sign due to the effect of the transposition. The near-end crosstalk at point A for the length $2D$ will, therefore, be the sum of the values for lengths AB and BC. This sum is:

$$NK\frac{1 - \epsilon^{-2\gamma D}}{2\gamma}(1 - \epsilon^{-2\gamma D}).$$

This quantity divided by NK is the type unbalance for the length $2D$ of Fig. 14. If D is only the length of a short segment the above expression is about equal to $NKD(2\gamma D)$.

Similarly the near-end crosstalk at point A for a length $3D$ will be:

$$NK\frac{1 - \epsilon^{-2\gamma D}}{2\gamma}(1 - \epsilon^{-2\gamma D} \mp \epsilon^{-4\gamma D})$$

and the type unbalance is this quantity divided by NK. For a length $4D$ the quantity in the parentheses becomes $(1 - \epsilon^{-2\gamma D} \mp \epsilon^{-4\gamma D} \mp \epsilon^{-6\gamma D})$, etc. The sign of each term in the parentheses is determined by the arrangement of "relative" transpositions, i.e., those at points where only one of the two circuits is transposed. Each term corresponds to a length D. The transposition at the start of the second length (at point B of Fig. 14) reverses the sign of the term for the second length and also the signs for the following lengths until another transposition is reached which makes the next sign plus, etc.

A practical open-wire line is divided into a series of "transposition sections" of eight miles or less. In each section the crosstalk between any two circuits is approximately balanced out by means of transpositions. A main purpose of this division into sections is to provide suitable points for circuits to drop off the line. A circuit on the line for a part of a section may have more crosstalk to a through circuit than if the parallel extended for the whole section since coupling in

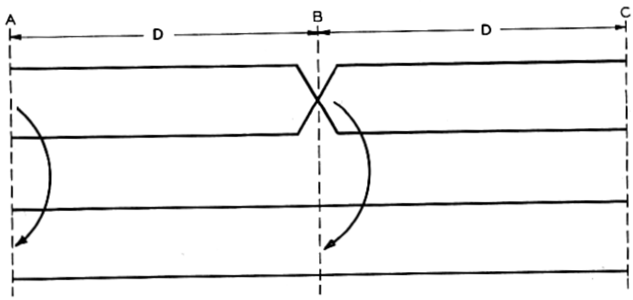

Fig. 14—Near-end crosstalk in length $2D$ between circuits a and b with circuit a transposed in the middle.

the last part of the section may tend to subtract from the coupling in the first part. The ends of sections are, therefore, the most suitable points for circuits to leave or enter the line. Ideally, the sections in a line should all be alike as regards length and transposition arrangements since this makes it practicable to so design the transpositions that residual crosstalk in one section tends to cancel that in another section. Practically, the sections vary in length and, therefore, in the transposition arrangements because the ends of some of the sections must fall at particular "points of discontinuity" determined by branching circuits and by requirements for balance against induction from power circuits.

In designing the transposition sections, type unbalances are computed for the section lengths of eight miles or less. For such lengths, the general method of computing type unbalances may be simplified. The general method involves the vector propagation constant γ. For a length as short as a single transposition section, attenuation can, ordinarily, be neglected. Therefore, in the type unbalance formulas γ can be replaced by $j\beta$ which greatly simplifies the computations.

Since attenuation can be neglected, the type unbalance for a transposition section depends only on the line angle βD. Since β increases practically directly with frequency, a plot of type unbalance against βD indicates the variation of type unbalance with frequency for a fixed length or the variation with length for a fixed frequency. It is convenient to plot the product of type unbalance and frequency (in kilocycles) since this product multiplied by the crosstalk coefficient gives the crosstalk. Two such plots for near-end type unbalance times frequency are shown on Fig. 15A. The plot marked P is for the condition of two circuits non-transposed or transposed alike. The plot marked O is for the same arrangement except for one relative transposition at the midpoint of the parallel.[4] The figure has a frequency scale corresponding to a length of eight miles as well as the general βD scale in degrees.

It will be seen that, for the case of no relative transpositions, the crosstalk varies directly with the frequency for only a short distance at the start of the curve. The effect of one relative transposition is to greatly reduce the crosstalk for small values of βL. For larger values the crosstalk is increased. It may be noted that the minimum values shown on the curves are somewhat in error since attenuation was neglected.

The minimum values in the P curve are due to "natural transpositions" in the non-transposed circuits. When the line angle is 180 degrees the crosstalk at the near-end of the disturbed circuit due to the second half of the line is just 180 degrees out of phase with the crosstalk due to the first half. This reversal in phase is due to the phase change accompanying the propagation of current to the midpoint and back. The total crosstalk due to both halves of the line lengths is the same as if the crosstalk coupling in the second half were translated to the near-end and the parallel without phase change but one circuit was transposed at the mid-point. When the line angle is 360 degrees the "natural transpositions" are at the quarter points, etc.

The near-end crosstalk between any two circuits in a transposition section may be estimated by multiplying the crosstalk coefficient by values of type unbalance times frequency similar to those of Fig. 15A. The total crosstalk in a succession of similar transposition sections is calculated at any particular frequency by working out a factor similar to the type unbalance in order to obtain the relation between the crosstalk in many transposition sections and that in one section. In calculating this factor, attenuation cannot be neglected since long lengths of line are involved.

[4] Two circuits are relatively transposed by one transposition at a given point in the line. Transpositions in both circuits leave them relatively untransposed.

The method of computing type unbalances for far-end crosstalk will now be explained. As in the case of near-end crosstalk, the type unbalance is defined by expressing the far-end crosstalk between two long circuits as the product of the crosstalk coefficient, the frequency in kilocycles and the type unbalance.

Figure 15B indicates the periodic variation with frequency of the far-end crosstalk when type unbalance is controlling.

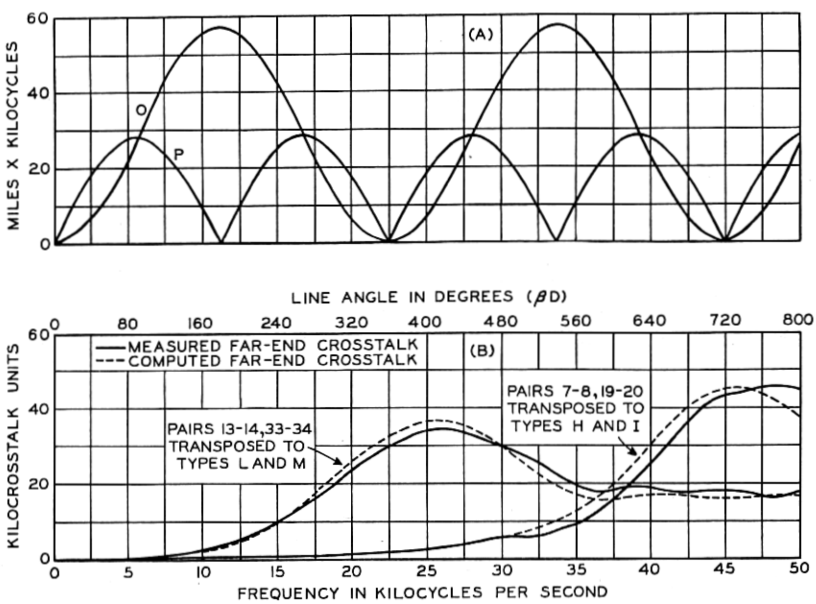

Fig. 15—Type unbalance and crosstalk vs. frequency and line angle in degrees. For Part (*B*), see Fig. 27*A* for wire configuration and Fig. 28 for transposition types.

In the case of near-end crosstalk, the method of computing the type unbalance neglected interaction crosstalk since, ordinarily, the transpositions needed to control transverse crosstalk make the interaction effect negligible. In the case of far-end crosstalk, the most important type of interaction crosstalk is included in calculations of type unbalances but another type of interaction crosstalk and the direct transverse crosstalk are neglected. The transpositions needed to properly suppress the important type of interaction crosstalk and the indirect transverse crosstalk ordinarily make the neglected types of crosstalk very small and the application of a more precise method of computing type unbalances for far-end crosstalk is not justified in practice.

The far-end type unbalance for a non-transposed part of a long parallel between two circuits will be computed first. Such a part of a parallel is indicated by length D of Fig. 16. For purposes of computation this length is divided up into a number of short segments each of length d. Considering the far-end crosstalk for two such segments at the start of the length D it will be seen from the discussion of crosstalk coefficients that transverse crosstalk in the length $2d$ will be

$$2FKd = 2(F_d + F_i)Kd.$$

In the above expression F is the far-end crosstalk coefficient, F_d being that part due to direct crosstalk and F_i that part due to indirect crosstalk.

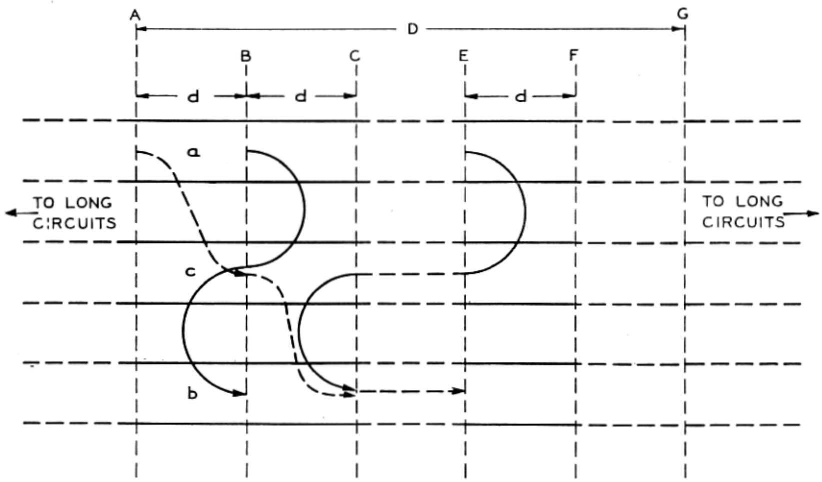

Fig. 16—Far-end crosstalk between untransposed circuits in length D.

The above expression relates to the output-to-output crosstalk. The input-to-output crosstalk is obtained by multiplying by the propagation factor $\epsilon^{-2\gamma d}$ to allow for propagation from A to C. This correction is usually made only when it is desired to obtain the input-to-output crosstalk between complete circuits and it is usually satisfactory to correct by using the attenuation factor and ignoring change in phase.

The total transverse output-to-output crosstalk in the length D is:

$$(F_d + F_i)KD.$$

This is about equal to F_iKD since F_d is ordinarily small compared to F_i.

Figure 16 indicates with a solid line the important type of interaction crosstalk between the first two segments by way of a representative tertiary circuit c. As discussed in the section on crosstalk coefficients and in Appendix A, the far-end crosstalk (output-to-output) of this interaction type will be

$$N_{ac}N_{cb}K^2d^2 10^{-6} = -2\gamma F_i K d^2 \text{ approximately.}$$

The interaction crosstalk as well as the transverse crosstalk is about proportional to the indirect coefficient F_i.

Each segment of the disturbing circuit will have a similar interaction crosstalk coupling with each preceding segment of the disturbed circuit. The interaction crosstalk between segment EF and segment BC is indicated on Fig. 16. The expression for this differs from the above expression in that the additional propagation distance from E to C and back must be allowed for. To get the total output-to-output far-end crosstalk it is necessary to sum up all these interaction crosstalk couplings between segments and to this sum add the total transverse crosstalk in length D.

This clumsy summation process may be avoided by letting d be an infinitesimal length and integrating between points A and G. This results in the following approximate expression for the output-to-output far-end crosstalk in the length D.

$$F_d KD + F_i KD + F_i K\left[\frac{1 - \epsilon^{-2\gamma D}}{2\gamma} - D\right].$$

This assumes the same propagation constant for the disturbing, disturbed and tertiary circuits. This approximation is justified for short lengths of, say, 10 miles or less.

The last term represents the interaction crosstalk and this term is negligible for small values of D. For larger values of D interaction crosstalk must be considered and it is convenient to rewrite the expression as follows:

$$F_d KD + F_i K\frac{1 - \epsilon^{-2\gamma D}}{2\gamma}.$$

The first term representing the direct crosstalk is negligible for values of D corresponding to a line angle of 90 degrees or less since F_d is ordinarily small compared with F_i and D is not large compared with $(1 - \epsilon^{-2\gamma D}/2\gamma)$. Therefore, direct crosstalk ordinarily may be neglected in computing far-end type unbalance. Another reason for neglecting direct crosstalk is that it is readily cancelled by a few relative transpositions while the remaining far-end crosstalk depends upon the transpositions in a complicated way, because the various interaction crosstalk couplings involve a variety of propagation distances and, therefore, have a variety of phase angles. If both circuits are transposed frequently but alike the direct crosstalk is not affected by the transpositions but it is ordinarily small compared with the indirect transverse crosstalk.

Figure 16 indicates by a dashed line another type of interaction crosstalk involving the product of two far-end crosstalk couplings. This effect can be neglected with practical arrangement of transpositions but may be important in the case of circuits having few transpositions or none at all.

In computing type unbalance the far-end crosstalk in an untransposed segment of line of length D may, therefore, be written as:

$$F_i K \frac{1 - \epsilon^{-2\gamma D}}{2\gamma} = FK \frac{1 - \epsilon^{-2\gamma D}}{2\gamma} \text{ approx.}$$

Since the magnitude of F_i is ordinarily about equal to that of F, the measured coefficient, it is usually satisfactory to use the latter value.

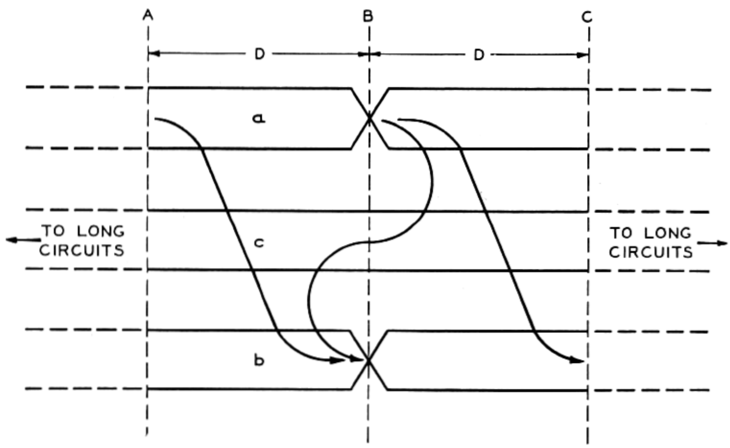

Fig. 17—Far-end crosstalk in length $2D$ between circuits a and b with each circuit transposed at the middle.

Having derived the above expression it is now possible to derive the far-end type unbalance for two transposed circuits. Figure 17 indicates a parallel between two long circuits. The type unbalance will be computed for a length $2D$ in which both circuits are transposed at the center. In the length $2D$ three far-end crosstalk paths must be considered, that is, the far-end crosstalk in length AB, that in length

BC and the important type of interaction crosstalk between length BC and length AB. The output-to-output crosstalk values only will be written for all these paths or, in other words, the effect of the propagation distance AC will not be considered in the expressions.

The far-end crosstalk for either length AB or BC is given by the above expression. Since *both* circuits are transposed at point B the far-end crosstalk values in the two lengths will add directly and their sum will be

$$2FK\frac{1-\epsilon^{-2\gamma D}}{2\gamma}.$$

Transmission from A to C through the crosstalk path in length AB is reversed in sign due to the transposition in circuit b at B. The output current of circuit a is also reversed in sign. In general, the output-to-output current ratio may or may not be reversed in sign depending on the transposition arrangement. It is convenient, however, to consider the first path as a reference and assign a plus sign to the crosstalk. Other paths are then assigned the proper relative phase angles.

As discussed in connection with Fig. 16, if the length D is very short the interaction crosstalk between the two segments may be written:

$$-2\gamma F_i KD^2 = -2\gamma FKD^2 \text{ approx.}$$

In practice the length D may be too long for this approximate expression in which case it is necessary to substitute for D in the above expression the value derived in connection with the discussion of the near-end crosstalk in a length D. In other words, D of the above expression should be replaced by

$$\frac{1-\epsilon^{-2\gamma D}}{2\gamma}.$$

With this substitution the interaction crosstalk between the two lengths becomes

$$-FK\frac{(1-\epsilon^{-2\gamma D})^2}{2\gamma}.$$

Transmission from A to C through this crosstalk path involves two transpositions and therefore the sign of the above expression is not reversed. Relative to the reference path through the crosstalk in length AB the sign should be reversed, however, and become plus. The total crosstalk in the length $2D$ is, therefore,

$$2FK\frac{1-\epsilon^{-2\gamma D}}{2\gamma}+FK\frac{(1-\epsilon^{-2\gamma D})^2}{2\gamma}=FK\frac{3-4\epsilon^{-2\gamma D}+\epsilon^{-4\gamma D}}{2\gamma}.$$

The latter expression divided by F is the frequency times the far-end type unbalance for the length $2D$. If one of the circuits were transposed at point B the crosstalk in length AB would be cancelled by that in length BC. The sign of the interaction crosstalk between the two lengths would be reversed and the expression would become

$$-FK\frac{(1-\epsilon^{-2\gamma D})^2}{2\gamma}.$$

If neither circuit were transposed at B, the far-end crosstalk would be that for a non-transposed length of $2D$ or:

$$FK\frac{1-\epsilon^{-4\gamma D}}{2\gamma}.$$

The frequency times the type unbalance values for the cases of one transposition and no transpositions are the same (in magnitude) as those derived for near-end crosstalk which were plotted (neglecting attenuation) as curves O and P on Fig. 15A.

If both circuits are transposed at B the near-end type unbalance remains the same as if there were no transpositions. The far-end type unbalance is radically altered, however. This is evident if the above equation is compared with that for the case of both circuits transposed.

This process of computing type unbalances may be extended from two equal lengths to any number of equal lengths. It is necessary to consider the interaction crosstalk between each length of the disturbing circuit and each preceding length of the disturbed circuit. The relative propagation distances through the various interaction crosstalk couplings must be taken account of.

Computations of far-end type unbalances are greatly simplified by assuming the same propagation constants for the disturbing, disturbed and tertiary circuits and by neglecting attenuation within a transposition section as in the case of near-end crosstalk. Since the tertiary circuit may be composed of any combination of wires on the line or of these wires and ground return, the propagation constant for a tertiary circuit may be somewhat different from that for the disturbing and disturbed circuits. This is particularly true of earth-return circuits, but these are of little practical importance due to their relatively high attenuation. All circuits not involving the earth have somewhere near the same speed of propagation but the tertiary circuits may differ greatly in attenuation constants.

For practical reasons a fair balance against crosstalk must be obtained in each transposition section (eight miles or less) and, as in the case of near-end crosstalk, type unbalances are calculated for the transposition arrangements which may exist in a single transposition section. Since the attenuation in a transposition section is not great, these calculations need not take into account differences in the attenuation constants of the various tertiary circuits. A long line has a series of transposition sections of various types and the total far-end crosstalk for any two circuits is a summation of the crosstalk values obtained from the type unbalances for the various sections plus interaction crosstalk between the various combinations of sections. With practical methods of transposition design, the transposition arrangements are so chosen that the interaction crosstalk between two sections is usually small compared with the far-end crosstalk in one section. A long line for the most part consists of a succession of similar sections with occasional sections of other types. Interaction crosstalk between dissimilar sections does not ordinarily contribute appreciably to the total far-end crosstalk. For the important case of a succession of similar sections interaction crosstalk between sections must be carefully considered since it may build up systematically and the total may be large compared with the summation for the far-end crosstalk values for the individual sections.

Serious interaction crosstalk between similar sections is guarded against by computing factors relating the far-end type unbalance in one section to that in various numbers of successive sections with various transposition arrangements at the junctions of sections. The factors actually computed are somewhat in error since they involve long distances and assume the same attenuation constants for all circuits. The errors are not sufficient, however, to prevent the factors from being a proper guide in avoiding systematic building up of interaction crosstalk between sections.

The above discussion assumes that the tertiary circuits are indefinitely extended or terminated to simulate their characteristic impedance. The tertiary circuits may not be terminated at the ends of a line since many of them are not used for transmission of speech or signals. Complete reflections of the crosstalk current in the tertiary circuits will, therefore, occur at their ends and these reflections somewhat modify the crosstalk currents in other circuits. This effect is important in a very short line since the reflected wave is again reflected at the distant end and at particular frequencies large changes in the tertiary crosstalk currents may occur due to multiple reflections. In a long line such multiple reflections are damped out and, in general, tertiary circuit reflection effects are not important.

If all the pairs on a line are transposed for the same maximum useful frequency, the transposed pairs will usually be relatively unimportant as tertiary circuits, that is, two pairs having small crosstalk between them usually contribute but little to the crosstalk between one of these pairs and any third pair. In some cases, however, this effect is important. On some lines certain pairs may be transposed for carrier operation and other circuits on the line for voice frequencies only. A combination of the two kinds of circuits may have large crosstalk between them at carrier frequencies and may contribute appreciably to the carrier frequency crosstalk between the pair transposed for carrier operation and some other pair also so transposed.

Far-end type unbalances which take account of transpositions in a tertiary circuit must, therefore, be calculated. This can be done by following the same general method discussed in connection with Fig. 17. From the discussion of coefficients it follows that the far-end coefficient for use in computing such a type unbalance will be:

$$\frac{-N_{ac}N_{cb}K}{2\gamma}10^{-6},$$

where N_{ac} and N_{cb} are the near-end crosstalk coefficients for the combination of disturbing circuit and tertiary circuit and the combination of tertiary circuit and disturbed circuit. Since these circuit combinations involve recognized transmission circuits, their near-end coefficients will be available since they must be measured or computed in order to compute the near-end crosstalk.

If a parallel between two circuits is divided into a large number of segments by transposition poles there is a wide variety of transposition arrangements which may be installed at these poles. It is, therefore, a complicated problem to devise charts and tables in reasonable numbers which will cover all the possible type unbalance values for the various transposition arrangements over a wide range of frequencies. This is particularly true in the case of far-end type unbalances since the type unbalance is altered by transposing both circuits at the same points and it is necessary to work out a type unbalance for each combination of transposition arrangements which may be used in two circuits. In the case of near-end crosstalk a number of different transposition arrangements will have the same type unbalance since only the relative transpositions need be considered.

The circuit capacity of a line may be increased by the use of phantom circuits (generally when carrier-frequency systems are not involved) which must, of course, be transposed to avoid noise and crosstalk. The crosstalk between phantom circuits may be calculated in a manner similar to that for pairs. The calculation of crosstalk between side circuits of the phantoms or between a side circuit and a phantom circuit is complicated by the fact that the phantom transpositions cause the side circuits to change pin positions. Near-end and far-end type unbalances have been computed, however, which take account of this "pin shift" effect of the phantom circuits. In general, the use of phantom circuits seriously limits the crosstalk reduction which may be obtained by transpositions. Phantom circuits are often uneconomic since they seriously restrict the number of carrier frequency channels which may be operated over a given pole line.

As indicated by Fig. 15A the values of type unbalance times frequency have marked maximum and minimum values when they are plotted against frequency or length. The maximum values are usually reduced by increasing the number of transpositions in a given length. When there are a number of circuits on the line it is usually necessary that the propagation of current between successive transposition poles does not change the phase by more than about five degrees. Since the phase change is about two degrees per mile per kilocycle the maximum transposition interval in miles is about $2.5/F$ where F is the frequency in kilocycles. This means .25 mile or 1300 feet at 10 kilocycles and .06 mile or 300 feet at 40 kilocycles.

It does not follow, however, that the least maximum value of type unbalance for a range of frequencies is obtained by using the greatest number of transpositions for a given number of transposition poles. This is illustrated by Fig. 15A which shows that the least maximum value is obtained with no transpositions rather than with one transposition. The total crosstalk current at a terminal is composed of numerous elements of various magnitudes and phase relations. The vector sum of these elements tends to be small at particular frequencies with no transpositions at all and it is important to preserve this tendency as much as possible when choosing an arrangement of transpositions. The vector sum of the elements can never be made zero since this would require that the circuits have no attenuation and infinite speed of propagation. This sum and, therefore, the type unbalances can be made very small, however, by choosing a suitable transposition arrangement and making the interval between transposition poles very small. In practice, the values of type unbalance times frequency for adjacent circuits are restricted to values much less than those of Fig. 15A.