Open-Wire Crosstalk *

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Effect of Constructional Irregularities

IF the cross-sectional dimensions of an open-wire line were exactly the same at all points and if the transpositions were located at exactly the theoretical points, the crosstalk could be reduced by huge ratios by choosing a suitable transposition arrangement and interval between the transposition poles.

Practically, however, the crosstalk reduction is limited by unavoidable irregularities in the spacing of the wires and of the transposition poles. There is no point in reducing the type unbalances by transposition design beyond the point where the constructional irregularities control the crosstalk.

Transposition Pole Spacing Irregularities

The following discussion covers the method of estimating the crosstalk due to irregularities in the spacing of transposition poles and the derivation of rules for limiting such irregularities. With practical methods of locating transposition poles, the effect of the pole spacing irregularities may ordinarily be calculated by considering only the transverse crosstalk. Special conditions for which attention must be paid to interaction crosstalk are discussed later. The simplest case, that of transverse far-end crosstalk due to pole spacing irregularities, will be discussed first.

A transposition section is divided into segments by transposition poles which in practice vary in number from four to 128. Each segment causes an element of crosstalk current at a circuit terminal and this element is about proportional to the segment length. For far-end crosstalk between similar circuits all these crosstalk current elements would add almost directly if there were no transpositions. The function of the transpositions is to reverse the phase of half the current elements. The segments corresponding to the reversed current elements may be called the minus segments. If the other half of the

* This is the second half of a paper which was begun in the January 1934 issue of the *Technical Journal*, giving 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.

segments are called the plus segments, the far-end crosstalk is proportional to the difference of the sum of the plus segments and the sum of the minus segments. This difference may be called the unbalanced length and the output-to-output far-end crosstalk is this length multiplied by the far-end coefficient and by the frequency.

If the sum of the plus segments equals the sum of the minus segments, the unbalanced length will be zero. The poles of a line are necessarily spaced somewhat irregularly but for a single circuit combination the unbalanced length could be made very small by carefully picking the transposition poles so as to keep the sums of the plus and minus segments about equal. This procedure is impractical, however, because many circuit combinations must be considered and because necessary line changes would prevent the maintenance of very low initial unbalanced lengths.

In practice, therefore, the segment lengths are allowed to deviate in a chance fashion from the mean segment length. The unbalanced length varies among the various circuit combinations depending on the arrangement of the transpositions which determines the order in which plus and minus segments occur. For any particular combination, the unbalanced length has a wide range of possible values and its sign is equally likely to be plus or minus.

In any transposition section, the length of any segment may deviate from the average segment length for that section. If the sum of the squares of all the deviations in each transposition section is known, the unbalanced length for a succession of transposition sections may be estimated, that is, the chance of the total unbalanced length lying in any range of values may be estimated.

Letting S_1^2 be the sum of the squares of the deviations for the first transposition section, etc., and letting R be the r.m.s. of all the possible values of the total unbalanced length in all the sections, the following approximate relation may be written:

$$R^2 = S_1^2 + S_2^2 + \cdots \text{ etc.}$$

The chance of exceeding the value R may then be computed. For example, there is about a one per cent chance that the total unbalanced length will exceed $2.6R$.

In making rules for locating transposition poles the first step is to determine a value for R. For example, if consideration of tolerable crosstalk coupling indicated that there should not be more than one per cent chance that the total unbalanced length in a 100-mile line would exceed one mile, then R, the r.m.s. of all possible values of the total unbalanced length, should not exceed 1/2.6 miles. Since R is calculated from the values of S for the individual transposition sections, a given permissible value of R may be obtained with various sets of values of S. It seems reasonable to determine individual values of S on the principle that a transposition section of length L_s should have the same probability of exceeding a given unbalanced length as any other section of the same length and that a section of length $2L_s$ should have the same probability as two sections of length L_s, etc. On this basis, the value of S^2 for any transposition section should be proportional to the section length L_s. This leads to the rule used in practice that for any transposition section S^2 should not exceed kL_s. If L_s and S are expressed in feet, a value of three for k is found suitable for practical use. The choice of a value for k will depend, of course, upon the cost of locating and maintaining transposition poles with various degrees of accuracy and upon the effect on the crosstalk of varying the value of k.

The above rule permits a large deviation at one point in a transposition section if it is compensated by small deviations in the rest of the segments. For example, with 128 segments and a mean segment length of 260 feet, one long segment of 575 feet is permissible if the rest of the segments are 258 feet. The expression for the total unbalanced length in a succession of transposition sections assumed that the deviations varied from segment to segment in a truly random manner. The above example involves an unusual arrangement of the deviations. When there are a number of transposition sections in a line, such unusual arrangements of deviations in various sections do not have much effect on the probability that the total unbalanced length will exceed a given value.

The computation of near-end crosstalk due to pole spacing irregularities is a more complicated problem since the crosstalk elements resulting from the various segments vary in their magnitudes and phase relations because the various segments involve different propagation distances. It may be concluded, however, that the r.m.s. value of the total unbalanced length in all the sections may be expressed as follows:

$$R^2 = S_1^2 + S_2^2A_1^2 + S_3^2A_2^2 + \cdots.$$

This differs from the expression for far-end crosstalk in that the values of S^2 for the second and succeeding transposition sections are multiplied by attenuation factors. The attenuation factor A_1^2 corresponds to propagation through the first section to the second section and back again. The other attenuation factors are similarly defined. The above expression neglects attenuation within any particular transposition section since this is ordinarily small. It also assumes that the rule for locating transposition poles, that is, that S^2 should not exceed kL_2, is applied for lengths having only negligible attenuation.

In making estimates of R in connection with transposition design work, it is assumed that all the segments are nominally the same length, D, and that r is the r.m.s. value of the deviations of the segments. Since r^2 equals S^2 divided by the number of segments in length L_s, r^2 should not exceed kD. R^2 may be expressed approximately in terms of r^2 as follows:

$$R^2 = r^2 \frac{1 - \epsilon^{-4\alpha L}}{1 - \epsilon^{-4\alpha D}},$$

where R and r are expressed in the same units, L is the length of the line in miles, α is the attenuation constant per mile, and D is the segment length in miles. If the line loss is 6 db or more the expression is nearly equal to:

$$R^2 = \frac{r^2}{.46Da},$$

where a is the line loss in db per mile and D is the segment length in miles. This assumes $4\alpha D$ is small compared to unity which is usually the case.

The chance that the total unbalanced length will exceed about $2.1R$ is estimated at 1 per cent.

For far-end crosstalk (output-to-output) the same assumption as to nominal segment length leads to the expression:

$$\frac{R}{r} = \sqrt{\frac{L}{D}}.$$

The general expressions given for R^2 suggest that a very long segment might be permitted at some point in the line if the deviations of the segments were properly restricted in other parts of the line. The expressions given for far-end and near-end values of R^2 were

$$R^2 = S_1^2 + S_2^2 + S_3^2 + \cdots,$$

$$R^2 = S_1^2 + S_2^2 A_1^2 + S_3^2 A_2^2 + \cdots.$$

If a very long segment at some point, such as a river crossing, were permitted, this would increase the sum of the squares of the deviations for some transposition section. For example S_3 might be abnormally large. R^2 could be kept at some assigned value by limiting S_1^2, S_2^2, etc. This procedure is not considered good practice because of the difficulty of maintaining some parts of the line with very small deviations of the segments from their nominal lengths.

A very long segment has another effect on near-end crosstalk not indicated by the above discussion. If there were no deviations in any of the segments, the near-end crosstalk would be the vector sum of a number of current elements of various magnitudes and phase angles and the sum would be small due to a proper choice of these magnitudes and angles in designing the transpositions. If a segment deviates from its normal length, the magnitude of the crosstalk due to the segment changes and the phase angle also changes. The phase angles of the crosstalk values due to succeeding segments are also changed since they must be propagated through the segment in question. For ordinary deviations in segment lengths these effects on the phase angles may be neglected.

Since transverse crosstalk is independent of transpositions occurring in both circuits at the same point, it would appear from the above discussion that the location of such transpositions need not be accurate. This is not ordinarily a question of practical importance. If some circuit combinations have both circuits transposed at a certain transposition pole there will usually be other combinations which have relative transpositions at this pole. The transposition pole is of importance, therefore, in connection with the latter combinations and the same accuracy of location is required for all transposition poles. A question of practical importance, however, is whether the above rules for locating transposition poles properly limit the interaction crosstalk. This is affected by transpositions in both circuits at the same pole as well as by relative transpositions. In the following discussion of this matter it is concluded that the effect of transposition pole spacing irregularities on interaction crosstalk may be ignored at frequencies now used for carrier operation.

The effect of deviations in segment length on interaction crosstalk is indicated by Fig. 18. This figure indicates a short part of a parallel between two long circuits a and b. A representative tertiary circuit c is also shown. The transposition arrangements are like those of Fig. 9B. In connection with the latter figure it was shown that the interaction crosstalk would be very small if all segments had the same length d. On Fig. 18, D is used to indicate the normal segment length and the deviation of two segments from D is indicated by d. Since the length AC equals the length CF, these deviations have no effect on the transverse crosstalk which is controlled by the transposition at C. The deviations affect the interaction crosstalk between the length CF and length AC.

The circuit a has near-end crosstalk coupling with circuit c in the length CF. This effect is normally practically suppressed by the transposition in a at E. Due to the deviation d of segment CE, the near-end crosstalk between a and c in length CF will not be suppressed but will be proportional to d. There will likewise be near-end crosstalk between c and b in the length AC proportional to d. The two deviations, therefore, introduce interaction crosstalk practically proportional to d^2.

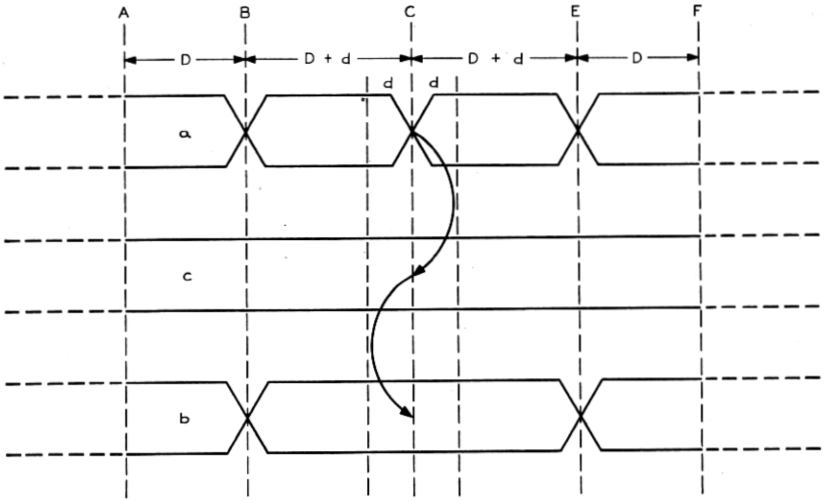

Fig. 18—The effect of deviations in segment length on interaction crosstalk.

Since there will be small deviations in numerous other segments of circuit b, the deviation d in circuit a will introduce numerous other interaction crosstalk paths similar to that discussed above. The r.m.s. value of the total interaction crosstalk caused by deviations in segment lengths may be roughly estimated as follows:

$$\frac{2FK\gamma r^2\sqrt{\frac{L}{D}}}{\sqrt{.46aD}} = \frac{.1FK^2r^2\sqrt{\frac{L}{D}}}{\sqrt{aD}},$$

where r is the r.m.s. deviation, L is the line length, D is the nominal segment length and a is the line loss in db per mile, all distances being expressed in miles. The above expression varies about as the 1.75 power of frequency and as the square of r. The corresponding expression for transverse crosstalk, i.e., $FKr\sqrt{\frac{L}{D}}$ varies as the first power of frequency and of r. It follows that, if the rules for accuracy of transposition pole spacing are relaxed or the maximum frequency is raised, the effect of pole spacing on interaction crosstalk increases more rapidly than the effect on the transverse crosstalk.

For the range of frequencies and accuracy of pole spacing used in practice, it has been found that the effect of pole spacing irregularity on interaction crosstalk is not controlling. This is indicated by Fig. 19

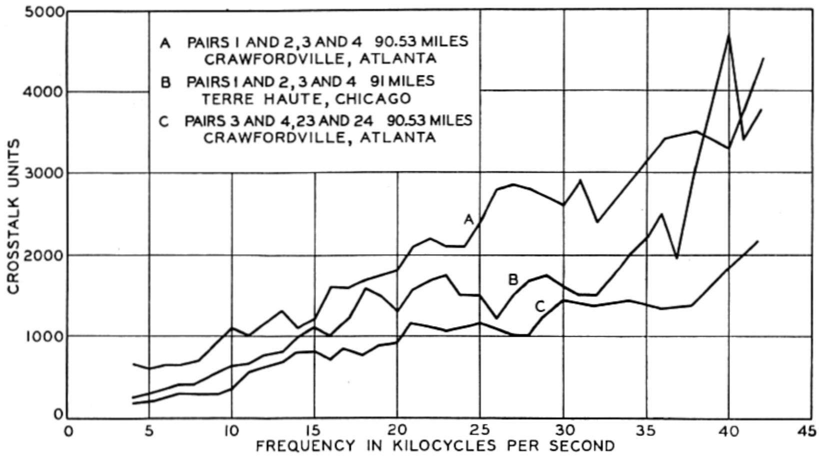

Fig. 19—Far-end crosstalk caused by pole spacing irregularities.

which shows some measurements of output-to-output far-end crosstalk between long circuits having transposition arrangements designed to make the crosstalk due to type unbalance small compared to that due to irregularities. The curves are about linear with frequency as would be predicted if the effect of the pole spacing irregularities (and wire spacing irregularities) on the interaction crosstalk is neglected. For these particular curves, a knowledge of the pole spacing indicated that pole spacing rather than wire spacing irregularities were controlling in causing crosstalk.

The above discussion assumes that a transposition section is divided by the transposition poles into segments all of the same nominal length. It is sometimes economical to use segments of different nominal lengths in the same transposition section. If the variation among the segment lengths is consistent rather than accidental it may be allowed for in the design of the transpositions.

In practice, segments of different lengths are used in the same transposition section when it is desired to adapt for multi-channel carrier frequency operation a few pairs on a line already having many pairs transposed for voice-frequency operation. Such lines often have existing transposition poles nominally spaced ten spans apart while for the pairs retransposed for carrier operation it is necessary to space the transposition poles about two spans apart. In such cases the cost of the carrier channels is appreciably increased if uniform spacing between the new transposition poles is used. The transpositions in the pairs retransposed for carrier operation must be coordinated with the transpositions in the other circuits and it is necessary, therefore, either to divide the ten spans into four approximately equal parts with consequent expense in setting new poles at the quarter points or to retranspose all the circuits on the line.

To avoid either of these expensive procedures, the new transposition poles are nominally located in the manner indicated by Fig. 20. This

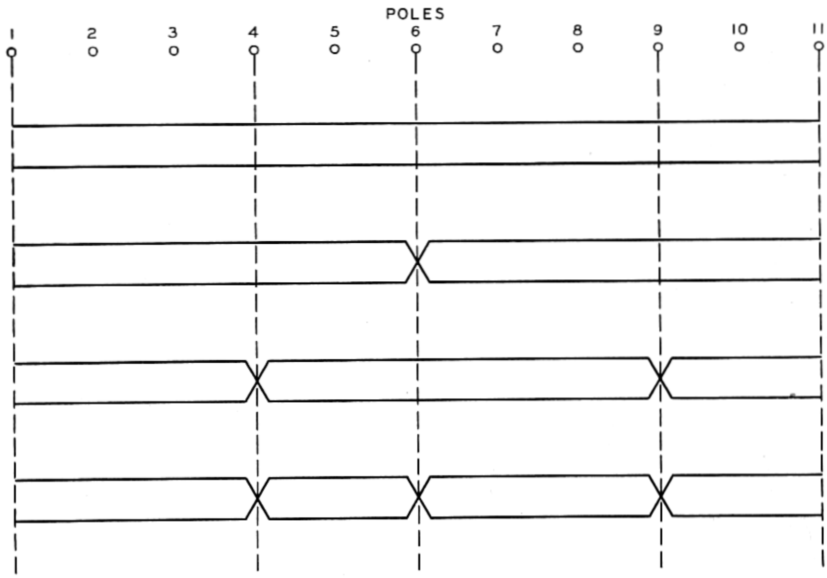

Fig. 20—Location of extra transpositions in a ten-span segment of line.

figure shows ten-pole spans subdivided into four parts in order to create three additional transposition poles. The figure indicates the location of the new transposition poles and the possible methods of transposing at these new poles. For some of the circuit combinations the crosstalk within the ten-span interval is considerably greater than if the four segments were equal in length. In each other ten-span interval the crosstalk is likewise increased by a similar inequality in segment length. Since all ten-span intervals are nominally alike, considerable crosstalk reduction may be obtained by properly designed transpositions located at the junctions of these intervals.

The use of segments of different lengths inherently decreases the effectiveness of the transpositions in reducing crosstalk and adds to the complexity of the transposition design problem. Uniform segments are therefore used except in special circumstances.

Wire Spacing Irregularities

In the past there has been a tendency to permit wire spacing irregularities in order to reduce the cost of construction and maintenance. For example, "H fixture" crossarms formerly had special wire spacing to permit the two poles to pass between pairs of wires and thus reduce the length of the arms. Another example is that of resetting a pole with a rotted base and reducing the spacing between crossarms to get clearance between wires and ground. The development of repeatered circuits and carrier current operation has increased the seriousness of the crosstalk resulting from such irregularities and made such practices generally undesirable.

There are, of course, unavoidable irregularities in wire spacing due to variations in dimensions of crossarms, insulators and pole line hardware and warping of crossarms. Corners and hills are other causes since the crossarms at a corner and the poles on a hill are not at right-angles to the direction of the wires. The most important unavoidable spacing irregularity is, however, due to variations in wire sag. Of recent years, limits have been set on wire sag deviations to insure that this effect is properly limited during construction. The main criterion adopted has been the difference in sag of the two wires of a pair. This difference is a rough measure of the crosstalk increment due to variations of the sag from normal. The crosstalk between two pairs in a given span will be abnormal if the two pairs have different sags even if there is no difference in sag for the two wires of a pair. The crosstalk is usually more nearly normal, however, than in the case of two pairs having the same average sag but different sags for the two wires of a pair. As far as practicable, all pairs are sagged alike in a given span.

The crosstalk between two pairs due to sag differences is computed much like that due to pole spacing irregularities. The change in crosstalk due to a known pole spacing deviation may, however, be computed from the crosstalk coefficient while the change in crosstalk due to a sag deviation is not related to the crosstalk coefficient in any simple way. Two methods have been used to obtain constants for calculation.

With the first method, crosstalk measurements were made on a long line (about 100 miles) having small pole spacing and type unbalance crosstalk. The r.m.s. of a number of crosstalk measurements was determined for each particular type of pair combination, for example, for horizontally adjacent pairs. The r.m.s. of the sag differences in a representative number of spans was also determined for the two pairs of each type of combination. The two r.m.s. values for any particular type of pair combination were called R and r. The ratio of R to r gave a constant k for estimating R from a known value of r and for L_0, the particular length of line tested. For other line lengths, R is estimated from the expression $R = kr\sqrt{\frac{L}{L_0}}$. Having computed R, the chance of the crosstalk for any pair combination in a long line lying in a given range may be estimated by probability methods.

The second method of studying sag differences is more precise although much more laborious. The change in crosstalk due to introducing sag differences in but two spans is determined. The poles are specially guyed to make it possible to adjust all the wires in these spans to have practically the same sag. Turnbuckles are installed at the ends of the two-span interval for this purpose. At the center pole the wires were supported so as to slip readily and equalize the sag in the two spans.

The phase and magnitude of the crosstalk is first measured for all pair combinations with all wires at normal sag. The wires are terminated in the same way as in the measurements of crosstalk coefficients. From sag measurements on actual lines, a set of unequal sag values for *all* the wires is then selected by probability methods and the crosstalk remeasured. The vector difference between the values of crosstalk before and after introducing unequal sags is then determined. This process is repeated a large number of times in order to cover the range of sag conditions encountered in practice. An r.m.s. value of the change in crosstalk due to sag difference is then determined for each pair combination and related to the r.m.s. sag difference per pair. This permits the probable crosstalk in a long line to be estimated and the importance of sag difference crosstalk to be determined. The two methods of study were found to be in general agreement. The second method has been extensively used to study proposed new wire configurations.

Drop Bracket Transpositions

An ideal transposition would cross the two sides of a circuit in an infinitesimally small distance, there being no displacement of the wires from their normal positions on either side of the transposition. The point-type transposition indicated by Fig. 21 is close enough to the ideal for practical purposes. Its deviation from the ideal requires little consideration in transposition design. To avoid cutting the wires, one wire is raised about 3/4 inch and the other lowered this amount at the transposition point. The drop bracket transposition

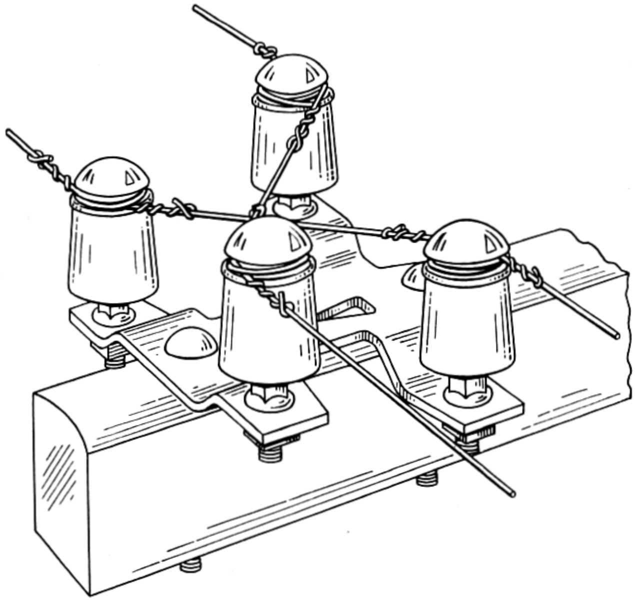

Fig. 21—Point-type transposition.

illustrated by Fig. 22 is considerably cheaper but the displacement of the wires is much greater. The effect of this displacement is important and must be especially considered in transposition design.

If all the spans adjacent to a drop bracket were of the same length

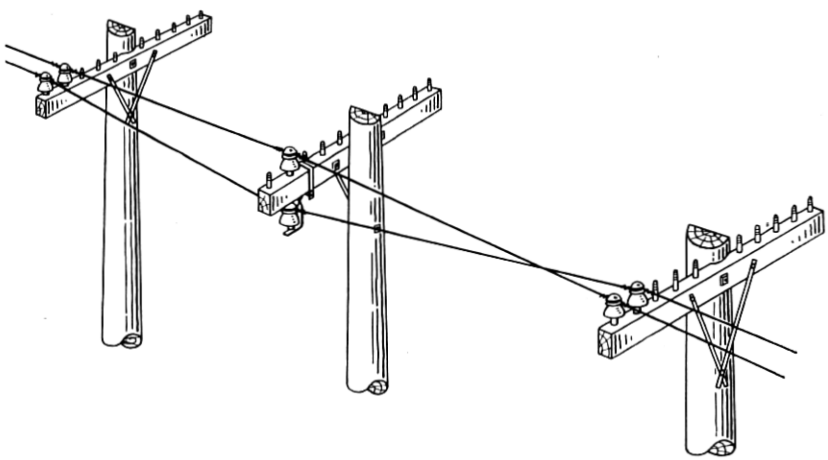

Fig. 22—Drop-bracket transposition.

and all wires could be kept under the same tension, the effect of drop brackets on crosstalk would be consistent and could, theoretically, be made negligible by a suitable transposition design.

There is, however, an accidental crosstalk effect. This effect is partly due to the fact that it is more difficult to avoid deviations from normal sag in the spans adjacent to drop brackets than in normal spans. The main effect, however, is thought to be due to inequalities in the lengths of the spans adjacent to drop brackets.

The crosstalk in such a span is very nearly proportional to the length of the span times a constant or "equivalent crosstalk coefficient." The usual crosstalk coefficient can not be used because the wires are not parallel.

Fig. 23-A indicates two long circuits, one circuit being transposed on drop brackets at the first and third quarter points of the short length D. The lengths of the spans adjacent to the drop bracket transpositions are indicated by d_1 to d_4. The equivalent far-end crosstalk coefficient for the span preceding a transposition bracket is F_1 and that for the span following the bracket is F_2. (F_1 and F_2 are usually quite different.) The total far-end crosstalk (output-to-output) due to the four spans is (very nearly):

$$K(F_1d_1 - F_2d_2 - F_1d_3 + F_2d_4),$$

where K is the frequency in kilocycles.

If the four spans were equal the crosstalk would be zero (very nearly). The actual value of the crosstalk is a matter of chance since the deviations of the four spans from the normal length are a matter of chance. These deviations cause a chance increase in the near-end crosstalk as well as in the far-end crosstalk.

This effect has been studied experimentally by using transposition designs which suppressed the consistent effect. The pole spacing effect was minimized by using very accurate spacing. The wire sag effect was allowed for by comparing similar pair combinations transposed alike except that dead-ended point transpositions were compared with drop bracket transposition. Due to the great number of transpositions necessary at carrier frequencies it was found that the accidental drop bracket effect was important at these frequencies. In recent years, point-type transpositions have been extensively used on lines transposed for long-haul carrier systems.

When, for economic reasons, a transposition system is designed for use with drop bracket transpositions, the consistent crosstalk effect must be considered in the transposition design. The equivalent crosstalk per mile for a span adjacent to a drop bracket must be determined for each pair combination. Approximate methods of computation have been worked out for doing this and checked against measurements. The computations are involved in connection with far-end crosstalk since the "tertiary effect" is controlling. Since the

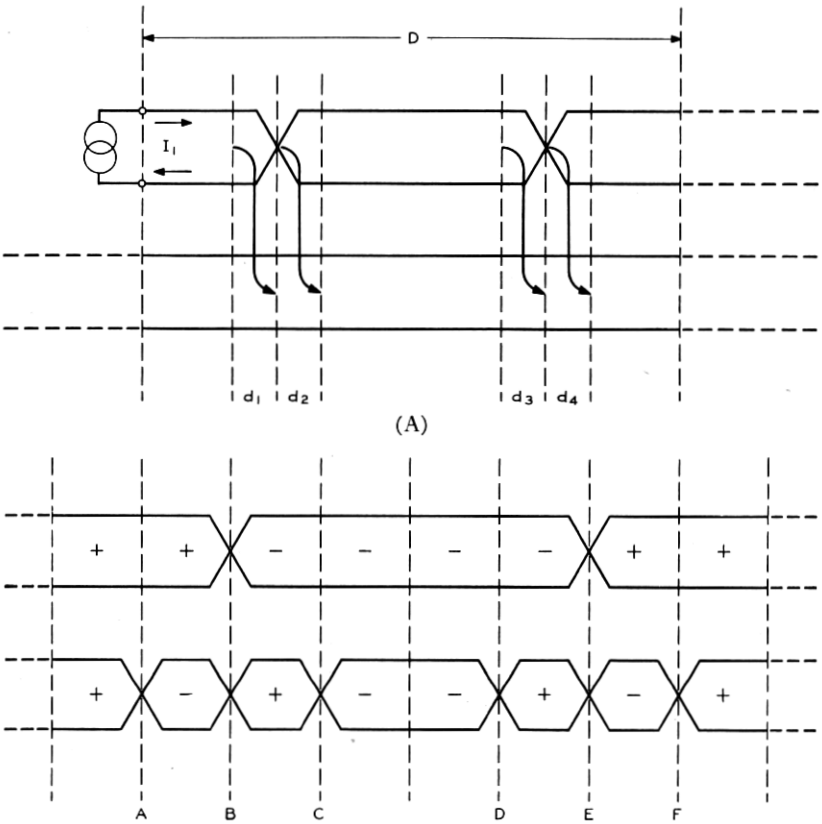

Fig. 23—Effect of drop brackets on crosstalk.

summation of crosstalk due to drop brackets is a consistent effect, "drop bracket type unbalances" can be worked out and used in transposition design. This matter is so complicated, however, that the practical method of design is to first practically ignore the drop bracket effect and then check the design to determine whether this effect has been properly suppressed.

Certain rules are adopted, however, to ensure that the transposition arrangements are properly chosen to avoid the larger drop bracket effects. Fig. 23-B indicates an arrangement of transpositions for two pairs in a short length of line which, with point transpositions, would have very low crosstalk. At points B and E both circuits are transposed alike. With point transpositions the near-end crosstalk in the two spans adjacent to one of these pairs of transpositions would be $NK2d$, where d is the span length, N the near-end crosstalk coefficient and K the frequency in kilocycles. For drop bracket transposition the crosstalk would be $K(N_1 + N_2)d$ or a change of $K(N_1 + N_2 - 2N)d$.

The transpositions are so arranged that the crosstalk in the two spans at B tends to add to that in the two spans at E. With drop brackets at B and E the major crosstalk in this length of line would be twice the above change since the crosstalk with point transpositions is very small.

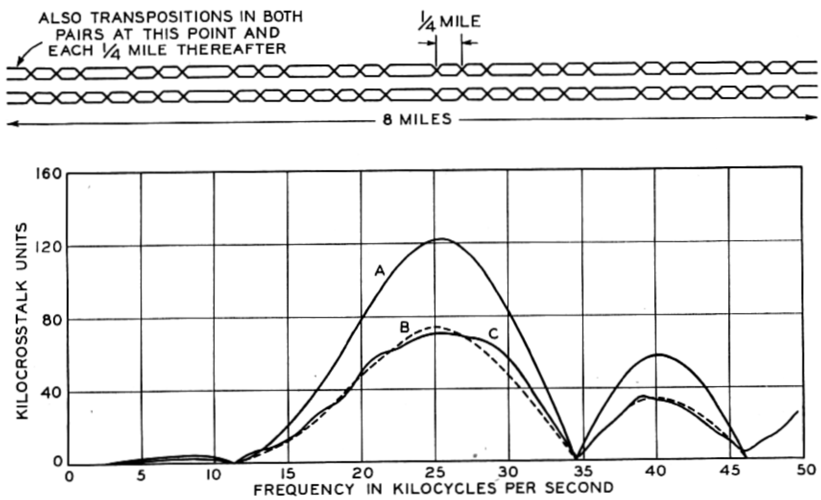

Fig. 24—Near-end crosstalk with and without drop brackets.

If the arrangement of Fig. 23-B is reiterated in a long line, the total increase in the crosstalk due to drop brackets at such points as B and E may be marked. It may be noted that the crosstalk in the two spans at A tends to cancel the crosstalk in the two spans at C and likewise there is cancellation at D and F. Drop brackets may, therefore, be used at points A, C, D and F without a consistent increase in crosstalk. Arrangements like those at B and E of Fig. 23B should be avoided in transposition design involving drop brackets.

The change in the crosstalk due to drop brackets is not necessarily an increase. Fig. 24 shows an arrangement of transpositions in an eight-mile line and three crosstalk frequency curves. Curve A shows the calculated near-end crosstalk for ideal point transpositions. Curves *B* and *C* show the calculated and observed near-end crosstalk for drop bracket transpositions. The curves show that the drop bracket effect can be calculated quite accurately and that it may reduce the total crosstalk. In the general case, it is impractical to take much advantage of this reduction effect because a marked reduction for one combination of circuits is likely to result in an increase for some other combination and because a reduction of crosstalk in one part of the line may increase the vector sum of crosstalk elements from all parts of the line.

Wire Configurations

The crosstalk coefficients for the various pair combinations may be altered by changing the configuration of the wires. Therefore, the crosstalk for a given transposition design and a given accuracy of transposition pole spacing irregularity may also be altered. The crosstalk due to sag differences also depends on the wire configuration. It is important, therefore, to choose a configuration most desirable from the crosstalk standpoint. Such an optimum configuration requires the fewest transpositions and least accuracy of pole spacing for a given maximum frequency and given permissible values of crosstalk coupling.

Various "non-inductive" arrangements of wire configurations have been suggested and tested. Such arrangements may appear to have possibilities but their study to date has indicated that they are impracticable for more than a few pairs on a line.

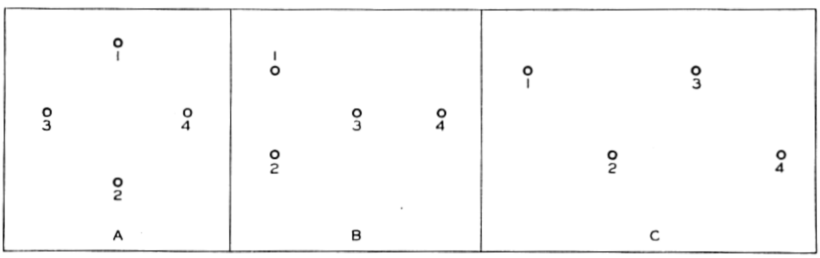

Fig. 25—"Non-inductive" arrangements for two pairs of wires.

Fig. 25 illustrates several suggested arrangements for two pairs. Arrangement *A* is often called a square phantom. If pair 1–2 is the disturber and there are equal and opposite currents in wires 1 and 2 there will be no voltages induced in either wire 3 or wire 4 because either of these wires is equally distant from wires 1 and 2. Since wires 1 and 2 are not equally distant from the ground, the currents in these wires may be not quite equal and opposite. As a result, voltages will be induced in wires 3 and 4 but these will be equal and there will be no crosstalk current in pair 3–4. By the reciprocal theorem the crosstalk between the two pairs will also be zero when pair 3–4 is the disturber.

Arrangement *B* is nearly non-inductive. In this case if pair 1–2 is the disturber and the currents in the two wires are not quite equal and opposite due to the presence of the ground, unequal voltages will be induced in wires 3 and 4 and there will be a crosstalk current in this pair. This effect could be minimized by transposing both pairs at the same points. They would not require relative transpositions since equal and opposite currents in pair 1–2 will induce no voltage in either wire 3 or wire 4.

With pair 3–4 as the disturber, equal and opposite currents will result in equal voltages induced in wires 1 and 2. These voltages cause a phantom current in phantom 1–2/3–4. This phantom current will divide between wires 3 and 4 but can not induce unequal voltages in wires 1 and 2 because 1 and 2 are equally distant from either 3 or 4. The crosstalk coefficient is, therefore, zero both for the direct effect and for the indirect effect of the phantom. However, the indirect effect of the ground or other conductors is not zero and may require transpositions.

Arrangement *C* is non-inductive for direct crosstalk. It is not non-inductive in regard to the indirect effect of the phantom 1–2/3–4. Equal and opposite currents in pair 1–2 induce equal voltages in wires 3 and 4. The resulting equal phantom currents in wires 1 and 2 of phantom 1–2/3–4 will induce unequal voltages in pair 3–4.

When there are many pairs on a line it is not possible to make all combinations strictly non-inductive even for direct crosstalk. With perfect wire spacing the larger values of direct crosstalk per mile could be greatly reduced, however, and appreciable reductions could be obtained in the indirect effect which is usually controlling in far-end crosstalk.

Wire sag deviations must be considered, however. If a given number of "non-inductive" pairs are placed in the pole head area normally occupied by the same number of pairs with conventional configuration, the crosstalk due to sag deviations is likely to be more serious with the "non-inductive" pairs than with conventional pairs. For the same pole head area, the number of transpositions and, therefore, the "pole spacing" crosstalk could be reduced if non-inductive arrangements were used. The tests to date indicate, however, that the total crosstalk would not be reduced because of increased "sag difference" crosstalk.

The mechanical problem of supporting the wires of the "non-inductive" arrangements is considerable if serious increases in crossarm and hardware costs are not to be incurred. This objection seems at present to override the possible advantages of (1) fewer transpositions for a given pole head area and crosstalk result, or (2) fewer transpositions and lower crosstalk with a greater pole head area.

Another possibility is the use of non-parallel wires. It is possible to arrange two pairs of wires in such a way that they have a certain direct crosstalk per mile at one end of a span and the value at the other end of the span is about equal and opposite. The net direct crosstalk per mile integrated over the span is zero or small. An example of this is the barreled square formerly used abroad. Fig. 26 illustrates this arrange-

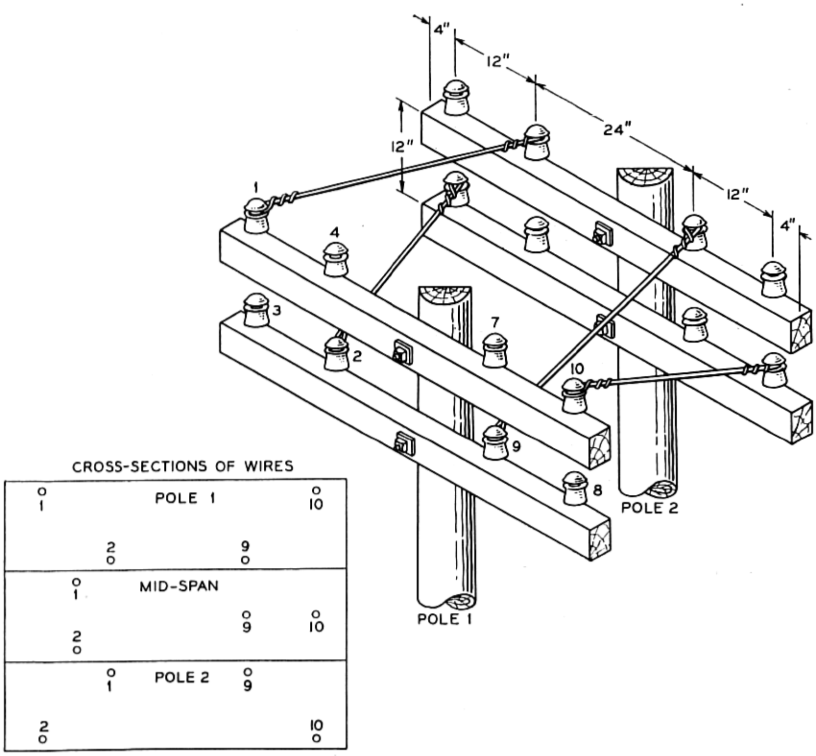

Fig. 26—Two pairs of wires in different barreled squares.

ment. The wires are arranged in groups of four, each four being arranged on the corners of a square. The two wires of a pair are on diagonally opposite corners of a square. Each pair is given a quarter turn in each span. For simplicity only two pairs in different four-wire groups and one span are shown. The two pairs shown are nearly "non-inductive" for direct crosstalk in this span.

Consideration has been given to applying this principle to a number of pairs in order to reduce the crosstalk coefficients. Since all the crosstalk coefficients could not be made very small, transpositions would be needed. The experience to date indicates that this method does not look attractive because it is not very effective in reducing the indirect crosstalk, the mechanics of transposing are difficult, the variations in sag are likely to be abnormal and the system is complicated.

There remains the simple method of improving the configuration of the wires in a given pole head area by reducing the spacing between the wires of a pair and increasing the spacing between wires of different pairs.

The crosstalk per mile between pairs is evidently reduced by this procedure since the two wires of a pair are approaching the ideal of being equally distant from every other conductor. The "sag difference crosstalk" is also reduced and higher frequencies may be used for a given crosstalk result. Fig. 27-A and Fig. 27-B indicate a 20-wire line with the wire spacing used in the past and also the configuration commonly used today on lines where heavy carrier development is involved. The spacing between the two wires of a pair has been reduced from 12 inches to 8 inches and the spacing between pairs correspondingly increased.

It was not possible to reduce the spacing of the pole pairs and for this reason they are unsuited for the higher carrier frequencies and it is sometimes uneconomic to string them. For such cases the crossarm indicated by Fig. 27-C may be considered. The 8-inch spacing of pairs is retained but the distance between pairs is further increased. With this last crossarm, phantom circuits are not superposed on the 8-inch pairs since their use results in greater crosstalk between the pairs and restricts the possibilities of multi-channel carrier operation.

The crossarm with 8-inch pairs and pole pairs may be used on lines where multi-channel carrier operation is not employed. In such cases, the 8-inch pairs may be phantomed. Since the average spacing between the side circuits of such a phantom is not reduced by the 8-inch spacing, the crosstalk between the phantom circuits is about the same as with the 12-inch pairs. The crosstalk from a side circuit into a phantom is somewhat reduced because of the reduced spacing of the pairs. For a given pole head area it does not appear practicable to devise a configuration which will result in marked reductions in the susceptibility of both phantoms and side circuits to crosstalk and

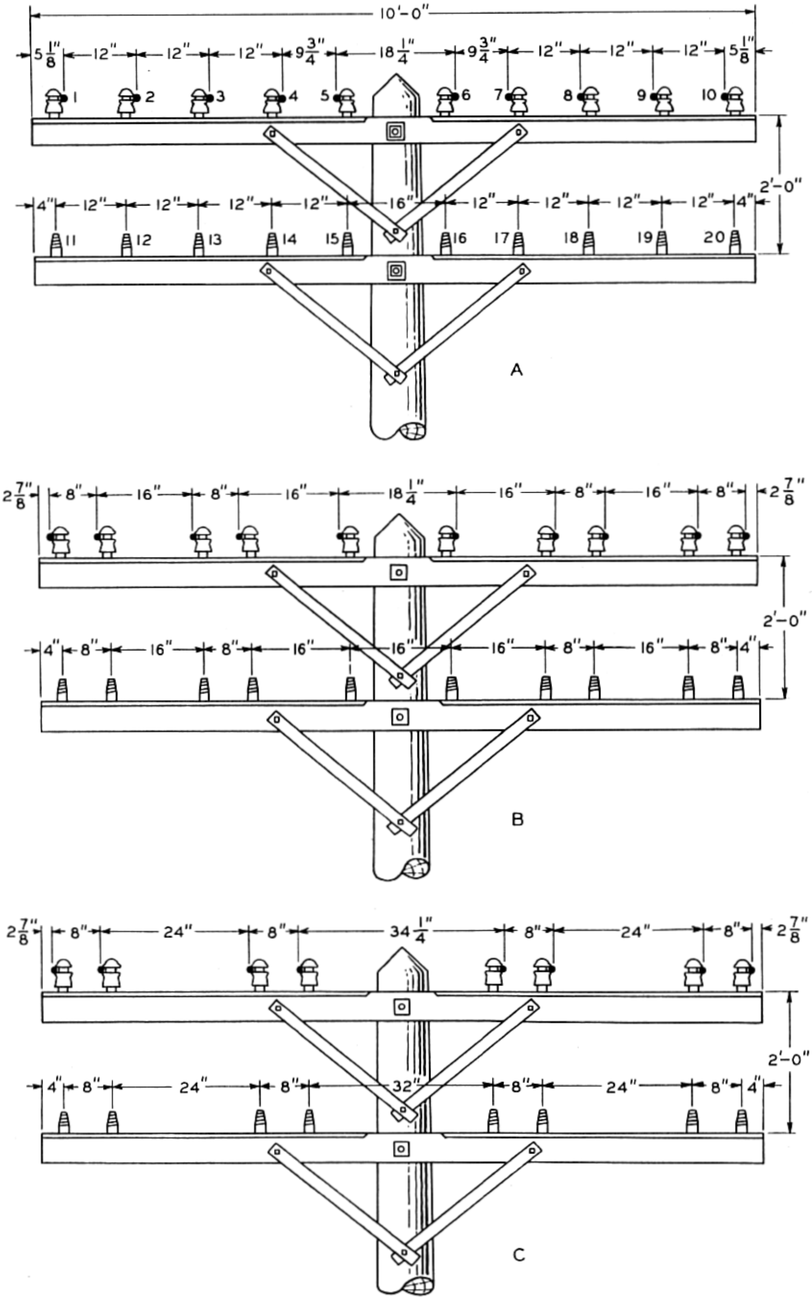

Fig. 27—Configurations of open-wire lines.

noise The "square phantom" indicated by A of Fig. 25 has theoretical possibilities but studies of the effect of wire spacing deviations make this arrangement appear impracticable.

The proposal to reduce the spacing of the wires of a pair from the historic value of 12 inches naturally raised the question of swinging contacts. However, extensive experience with 8-inch spacing has shown no appreciable increase in the number of wire contacts. This applies to lines where ordinarily the span length did not exceed about 150 feet. With long span crossings, crossarms were supported from steel strand at intervals of 260 feet or less.

The effectiveness of the reduction in wire spacing is indicated by the following table. The table shows the measured near-end and far-end crosstalk coefficients for important circuit combinations and for the two-pole head diagrams of Figs. 27-A and 27-B.

Crosstalk Per Mile Per Kilocycle—104-Mil Conductors

Pair Combination	Near-End Crosstalk		Far-End Crosstalk	
	12-Inch	8-Inch	12-Inch	8-Inch
1–2 to 3–4	974	439	74	34
3–4 to 7–8	133	47	77	15
1–2 to 11–12	653	326	66	30
1–2 to 13–14	40	18	58	24
3–4 to 13–14	549	288	155	69
1–2 to 21–22	163	78	35	16
1–2 to 23–24	55	28	43	17
3–4 to 23–24	107	55	75	36

General Transposition Design Methods

The preceding discussion will indicate that transposition design involves much more than consideration of the locations of the transpositions.

In practical design, the first step is to estimate the crosstalk due to unavoidable pole spacing and wire spacing irregularities for the configuration of wires under consideration and for a wide frequency range. This crosstalk represents the best that can be done with an ideal transposition design. It must be kept in mind that great precision is impracticable. The pole spacing of a line may change from time to time due to minor reroutings caused by highway changes, etc. The wire sag differences change with temperature and are affected by sleet.

If two long circuits are on adjacent or nearby pairs in one repeater section, they should, as far as practicable, be routed over non-adjacent pairs in other repeater sections in order to minimize the overall crosstalk between these two circuits. This crosstalk will usually be largely due to those parts of the parallel where the circuits are on adjacent or nearby pairs, since the pole line seldom has enough pairs to make it practicable to keep any two circuits far apart for a large proportion of the total parallel. It is important, therefore, to strive for the lowest possible crosstalk between adjacent or nearby pairs even though this requires permitting higher crosstalk between widely separated pairs than would otherwise be necessary.

For the adjacent or nearby pairs with naturally high crosstalk, limits on the type unbalance crosstalk are set which make this type of crosstalk small compared with that due to irregularities. Since the type unbalance crosstalk varies with frequency and, in general, increases with frequency, these limits are imposed only for the range of frequency which the line will be required to transmit. It is not advisable to go beyond this, since more severe limits require closer spacing of transpositions and the increased number of transpositions would make the "pole spacing" irregularity crosstalk larger. For the well-separated pairs with naturally lower crosstalk, the type unbalance crosstalk rather than the irregularity crosstalk may be allowed to control with the same idea in mind of requiring a minimum number of transposition points.

Fig. 28 indicates the method used generally in the Bell System for arranging transpositions with 32 transposition poles. The arrangements shown are called fundamental types. They are iterative, i.e., if the first two-interval length is transposed at the center, each following two-interval length is likewise transposed, etc. Various other arrangements called hybrid types are possible but in the long run there appears to be no advantage from their use except in the case of side circuits of phantoms. In this case the transposition pattern may change when the side circuit changes pin positions at a phantom transposition.

The fundamental types may be extended to involve 64, 128, 256, etc., transposition poles. Types involving 128 transposition poles are often used.

A long line, say 100 miles, is divided into short lengths called transposition sections. With the latest transposition designs, sections having 128, 64, 32, 16 and 8 transposition poles are provided. The nominal lengths of these sections vary from 6.4 to .25 mile. The purpose of these sections is to provide an approximate balance against crosstalk (and induction from power circuits) in short lengths and thus to allow for unavoidable discontinuities in the exposure between circuits such, for example, as points where circuits branch off the line. Transposition arrangements must be chosen for each circuit in each type of section to ensure this approximate crosstalk balance.

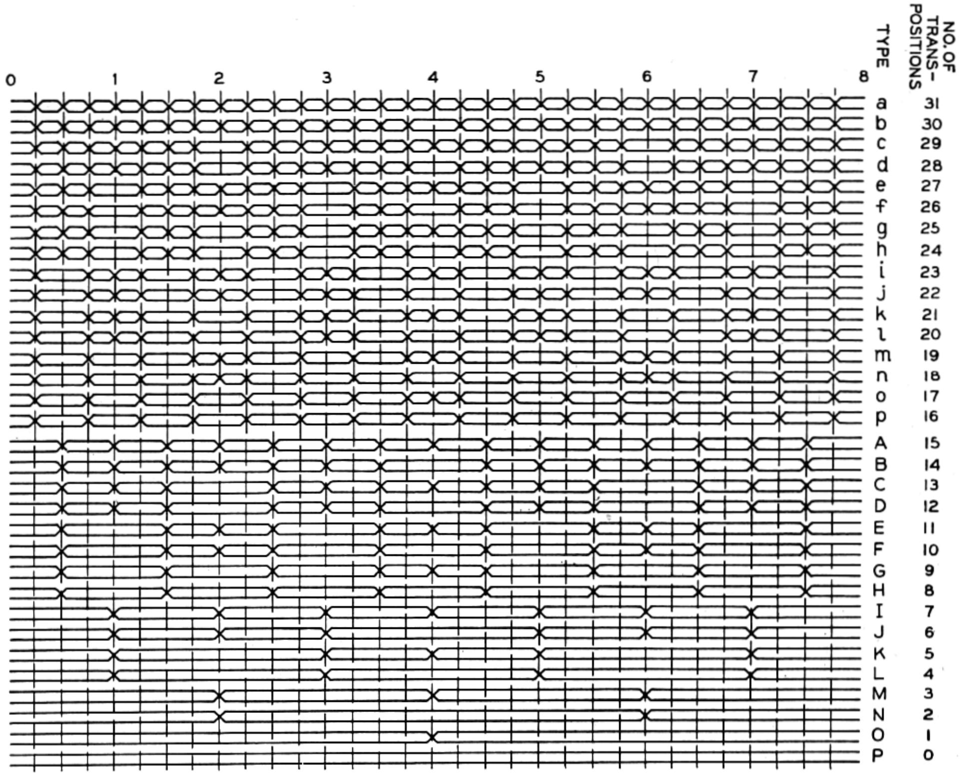

Fig. 28—Fundamental types for 32 transposition poles.

Certain lines have few, if any, discontinuities and a succession of the longest type of section is used. To improve the effectiveness of the transpositions, junction transpositions are used at the junctions of successive similar sections. For such lines it would be more effective to use longer transposition sections and not require that all circuits be approximately balanced in a short length. Such a special design would be impracticable, however, since it would be too inflexible in regard to circuit changes, etc.

In choosing the transposition arrangements for a section it must be kept in mind that the object is to meet certain crosstalk limits for a succession of sections considering both type unbalance and irregularity crosstalk. The method of procedure is discussed below.

Evolution of Transposition Designs

In designing transposition systems it must be kept in mind that much of the crosstalk is due to irregularities and is a matter of chance. Theoretically the crosstalk elements due to all of the various irregularities might chance to add directly. This is highly improbable and if the design were based on making this limiting condition satisfactory, the expense would be very great. Practically, therefore, the designs are based on exceeding a tolerable value a small percentage of the time. If, in practice, the tolerable value happens to be exceeded and this is not found to be due to an error in construction, the unfortunate adding up of crosstalk elements can be broken up by a different connection of circuits at the offices.

The tolerable values commonly chosen are 1000 crosstalk units (60 db) for open-wire carrier circuits and 1500 units (56 db) for voice-frequency open-wire circuits, which tend to have more line noise than cable or carrier circuits. These limits apply to the crosstalk between terminating test boards with the circuits worked at net losses of about 9 db.

Before proceeding with the design of the individual transposition sections which are but a few miles long, it is evidently necessary to determine what part of the overall limit can properly be assigned to an individual section. Assumptions must first be made as to typical and limiting lengths in which circuits are on the same pole line and in which adjacent or nearby circuits continue in this relation. A representative repeater layout must then be chosen. The repeater layout is very important, since the crosstalk in each repeater section is propagated to the circuit terminal and amplified or attenuated, depending on the arrangement of the repeaters. As a matter of fact, the layout of repeaters must be governed to a considerable extent by crosstalk considerations.

On the assumption that the relative magnitudes and phase relations of the crosstalk couplings in the various repeater sections are a matter of chance the tolerable crosstalk in a single repeater section can be estimated by the use of probability laws. Similarly the tolerable value for any part of the repeater section can be estimated. These probability methods apply very well to crosstalk due to irregularities. Type unbalance crosstalk is systematic, however, and in assigning tolerable values of type unbalance crosstalk in a transposition section, it is necessary to consider how the crosstalk values for various transposition sections may add up.

It is not likely that there will be systematic building up of type unbalance crosstalk in successive repeater sections and, therefore, the tolerable crosstalk per repeater section may be estimated by probability methods. The total of the irregularity crosstalk and the type unbalance crosstalk in a repeater section is a matter of chance and may be estimated from probability theory. Conversely, the part of the tolerable crosstalk which may be assigned to type unbalance crosstalk may be estimated. As noted above, the allowance for type unbalance crosstalk for adjacent or nearby pairs is usually made so small that irregularity crosstalk controls the total. The maximum permissible carrier frequency is, then, the frequency at which the irregularity crosstalk just reaches the tolerable value. Having determined tolerable values of type unbalance crosstalk for a repeater section for the various pair combinations, tolerable values for the individual transposition sections must be determined.

If a repeater section involves a number of different types of transposition sections it is not likely that there will be a systematic building up of type unbalance crosstalk. Factors are, therefore, worked out to relate the crosstalk in a succession of similar transposition sections to that in one section. Numerous factors are required since they depend upon the transpositions at the junctions of the sections. A study of such factors indicates values which it is reasonable to assign to an individual transposition section in order to avoid excessive type unbalance crosstalk in a complete repeater section.

In the case of a voice-frequency transposition system, both near-end and far-end type unbalance limits must be set. The far-end limits are usually easily met. In the case of a transposition system for carrier systems, far-end crosstalk is controlling and the far-end type unbalance limits are important. The "reflection crosstalk" previously discussed depends, however, on both the magnitude of the near-end crosstalk and on the impedance mismatches. Information on the degree to which it is practicable to reduce these mismatches must be available in order to set limits on near-end type unbalances at carrier frequencies.

Pairs used for carrier systems are usually also used for voice-frequency telephone systems and in designing transpositions for these pairs crosstalk limits suitable for both types of systems must be met. In practice, an existing line may have only a part of the pairs retransposed for carrier operation and in designing a system of transpositions for such retransposed pairs limits must be set for the crosstalk at voice frequencies between the retransposed pairs and the pairs not retransposed.

It has been the practice to transmit certain carrier telegraph frequencies in the opposite directions used for these frequencies in connection with carrier telephone, or, in some cases, program transmission circuits. At these frequencies near-end crosstalk limits must be set so as to limit the induced noise from the carrier telegraph.

When the type unbalance crosstalk limits are finally determined, the transposition designer must attempt to meet the requirements for all circuit combinations and all the transposition sections. It may be that the requirements can not be met and consideration must be given to modifications in the nature of the transmission systems. A vast amount of such preliminary transposition design work has been necessary in order to evolve the present transposition systems and transmission systems.

Such studies led to the development of non-phantomed circuits with 8-inch spacing since they indicated that multi-channel long-haul carrier operation on all pairs on a line was, in general, impracticable from the crosstalk standpoint with 12-inch phantomed pairs.

It may be noted that there are also difficulties in the crosstalk problem when 12-inch phantomed pairs are used for voice-frequency repeatered circuits. These circuits have a crosstalk advantage over carrier circuits in that the frequency is lower but they have an offsetting disadvantage in that they use the same frequency range in both directions. This makes the near-end crosstalk directly audible to the subscriber. As previously discussed the near-end crosstalk is inherently greater than the far-end crosstalk and, for this reason, practicable designs of multi-channel carrier systems do not allow near-end crosstalk to pass to the subscriber, the path being blocked by one-way amplifiers. While it takes fewer transpositions to control the type unbalance effects with voice-frequency transposition designs, for a given length of parallel the difficulties with crosstalk due to irregularities are about as great as with designs for multi-channel carrier operation.

The simple example of Fig. 29 illustrates the reasons for the difficulties with near-end crosstalk with the voice-frequency designs for 12-inch spaced pairs. It also illustrates the method of deducing the permissible crosstalk per repeater section as discussed above.

This figure indicates two paralleling repeatered circuits, each having six repeater sections of 10 db loss and five repeaters of 10 db gain. The net loss of each circuit is, therefore, 10 db. The near-end crosstalk values in the six sections are indicated by n_1 to n_6. The crosstalk coupling at A due to n_2 is just equal to n_2 since there is no net loss or gain in either circuit between A and B. There is also no net loss or gain between A and C, A and D, A and E or A and F. The total crosstalk coupling at A is, therefore, the vector sum of the six values

n_1 to n_6. If the crosstalk is due to irregularities the exact values of n can not be calculated but from the data collected on the crosstalk due to irregularities, the r.m.s. of all possible values may be estimated.

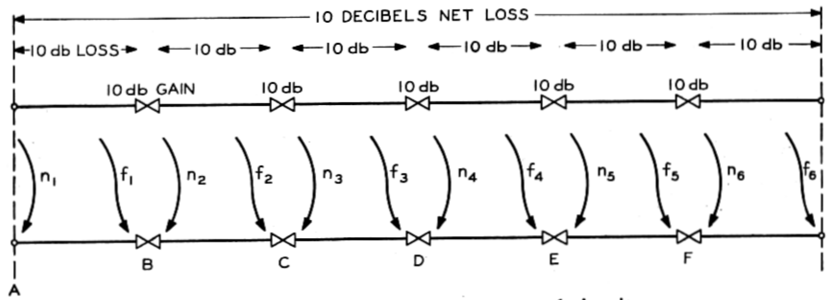

Fig. 29—Crosstalk between repeatered circuits.

Letting r_1 equal the r.m.s. value of n_1, etc., and using probability theory we may write:

$$R^2 = r_1^2 + r_2^2 + \cdots + r_6^2,$$

where R is the r.m.s. of all possible values of the near-end crosstalk at A. If $r_1 = r_2$, etc.

$$R = r\sqrt{6}.$$

The chance of the overall crosstalk deviating from R by any specified amount may be estimated by probability methods. It will be noted that the crosstalk in six repeater sections tends to be more severe than that in one section by $\sqrt{6}$ or, in other words, that the crosstalk varies as the square root of the length. If the use of repeaters were avoided by using more copper, for the same overall loss the crosstalk would be practically the same as with the repeatered circuits. *With the arrangement of repeaters shown* it is not the use of repeaters which causes the increase in crosstalk but rather the increase in circuit length without corresponding increase in circuit loss. For a given circuit length, circuit loss and wire size, other arrangements of repeaters may cause greater or less crosstalk.

If the repeaters of Fig. 29 are spaced farther apart, say 15 db instead of 10, there will be three line repeaters of 15 db gain each and terminal repeaters will be necessary to supply a terminal gain of 5 db in order to obtain a net loss of 10 db. The near-end crosstalk would be reduced by about $\sqrt{4} \div \sqrt{6}$ or 1.8 db because there are only four repeater sections but the terminal repeaters would amplify the near-end crosstalk by 5 db. The net increase would be 3.2 db. From the standpoint of near-end crosstalk, it is thus seen that close spacing between repeaters is very desirable.

In Fig. 29 the output-to-output far-end crosstalk in each repeater section is indicated by f_1 to f_6. The transmission path through any one of these crosstalk couplings is (for like circuits) a loss 10 db greater than the value of the coupling expressed as a db loss. With the repeater arrangement of the figure, the far-end crosstalk paths are attenuated by 10 db while the near-end crosstalk paths are not attenuated. Furthermore, the far-end crosstalk paths ordinarily introduce greater losses than the near-end paths. With greater spacing between repeaters, the near-end crosstalk is amplified but the far-end crosstalk (for like circuits) is still attenuated by the net loss of the circuits. At a given frequency the near-end crosstalk between such "two-wire" circuits is, therefore, much greater than the far-end crosstalk.

Review

Evidently the problem of keeping crosstalk between open-wire circuits within tolerable bounds is by no means a simple one. As we have seen, the work begins with consideration of complete circuits (telephone, program transmission or carrier telegraph) which may be hundreds or even thousands of miles long. The total crosstalk allowance for such long circuits must first be broken down into allowances for the various sections of line between repeaters and then into allowances for the individual transposition section, these individual sections ranging from less than 1/4 to about 6 miles in length.

Then bearing in mind that irregularities in pole spacing and in wire configuration set limits to crosstalk reduction which it is not practicable to overcome by transpositions, the crosstalk designer determines by computation whether, when considering these irregularity effects alone, the crosstalk requirements for the individual transposition sections can be met. If these requirements can not be met he must either have the general circuit layout altered so that, for example, the repeater gains will be more favorably disposed from the standpoint of crosstalk, or he must alter the pole head configuration so that the electrical separation between the circuits will be increased.

Having obtained an overall circuit layout and a configuration of the wires which makes it possible to attain the desired overall crosstalk results, the design of the transpositions proper is undertaken. In this work the transposition designer makes every effort to keep the number of transpositions at a minimum. He does this partly to save money but more particularly because he recognizes that more than enough transpositions do harm rather than good by increasing the number of pole spacing irregularities.

In dealing with the problems of crosstalk coupling between open-wire circuits, consideration must be given not only to the direct effect of one circuit on another but also to the indirect effect of the other circuits on the line. What happens is that the disturbing circuit crosstalks not only directly into the circuit under consideration but also into the group of other circuits and thence into the disturbed circuit. The name "tertiary" circuit has been given to this group of circuits although it is not in reality one circuit but rather any or all of the possible circuits which may be formed of the different wires. The system of transpositions must, therefore, not only substantially balance out the direct couplings between disturbing and disturbed circuits but must also substantially balance the couplings from the disturbing circuit into the "tertiary" circuit and from this "tertiary" circuit into the disturbed circuit.

Reflections of the electrical waves also add interest and complexity to the problem. Such reflections tend to increase crosstalk because the electrical waves which are changed in direction as a result of reflections crosstalk differently, and in many cases more severely, into neighboring circuits than do the waves traveling in the normal direction. The most important reflections occur at junctions between lines and office apparatus. The possibility of other reflections must also be considered, however, at intermediate points in the line which might be caused by inserted lengths of cable, change in spacing of wires, etc.

In working out the transposition designs, the fact that crosstalk between two paralleling circuits tends to manifest itself at both ends is of great importance. At the "near end" crosstalk coming from the disturbed circuit in a direction opposite to the transmission in the disturbing circuit must be considered. At the "far end" crosstalk coming in the same direction as the transmission in the disturbing circuit must be considered.

For telephone circuits which use the same path for transmission in both directions, the "near-end" crosstalk is considerably more severe than the "far-end" for two reasons: (1) The crosstalk per unit length of the paralleling circuits is greater; (2) the gains of the repeaters especially augment the "near-end" crosstalk. Voice-frequency open-wire telephone circuits have always been worked on this "one-path" basis and are good examples of circuits in which "near-end" crosstalk is controlling and must be given principal consideration in working out transposition designs.

In the case of carrier circuits, it was found early in the development that if these circuits were worked on a one-path basis, the crosstalk would be prohibitively great. Consequently, carrier circuits are now designed to operate on a two-path basis. Two separate bands of frequencies are set aside, each being restricted, by means of one-way amplifiers and electrical filters, to transmission in one direction only. Each telephone circuit is then made up of two oppositely directed channels, one in each frequency band. Thus, direct "near-end" crosstalk is kept from passing to the telephone subscribers. Consequently, the "near-end" type of crosstalk needs to be considered only with respect to that portion which arises from electrical waves reflected at discontinuities in the circuits, which effects have already been mentioned.

In practice a pole line may have some of the pairs very frequently transposed to make them suitable for carrier frequency operation and other pairs less frequently transposed and suitable only for voice-frequency operation. A system of transpositions must permit any arrangement of the two types of pairs which may be found economical for a given line and layout of circuits. Each pair must meet limits for near-end and far-end crosstalk to any other pair which may crosstalk into it in its frequency range. Pairs used only for voice frequencies are usually phantomed and transpositions must, of course, be designed for the phantom circuits as well as the side circuits. The design of a transposition system is, therefore, extraordinarily complicated and tedious and, to paraphrase the Gilbert and Sullivan policeman, "A transposer's lot is not a happy one."

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APPENDIX

Calculation of Crosstalk Coefficients

This appendix will first cover methods of calculating the coefficients of transverse crosstalk coupling. It is necessary to calculate both near-end and far-end crosstalk coefficients which involve both direct and indirect components of transverse crosstalk coupling. Coefficients for the direct and for the indirect components will be derived separately and then combined to obtain the total coefficients.

Ordinarily, the indirect effect cannot be readily computed with good accuracy and the total coefficients are usually measured. As previously noted, the method of computing the indirect effect can be used with fair accuracy, however, and it is useful in cases where measurements are impracticable.

The crosstalk between frequently transposed circuits may be calculated with the aid of the above coefficients of transverse coupling and in addition an "interaction crosstalk coefficient" relating to interaction crosstalk coupling of the most important type. The relation of this interaction coefficient to the far-end coefficient of transverse coupling is also discussed herein.

Direct Crosstalk Coefficients

Figure 30 indicates the definitions of the direct crosstalk coefficients used in computing the direct component of the transverse crosstalk coupling. This figure shows a thin transverse slice in a parallel between two long circuits a and b, the thickness of the slice being the

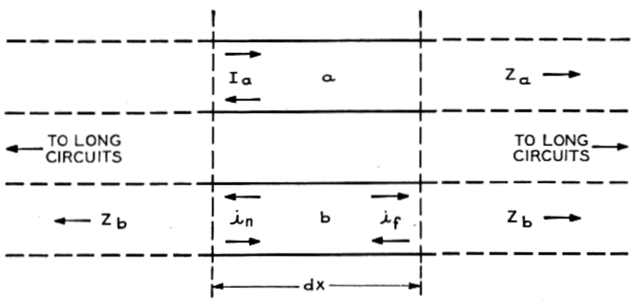

Fig. 30—Crosstalk in a single infinitesimal length.

infinitesimal length dx. Circuit a is energized from the left, the current entering dx being I_a. Propagation of I_a through dx results in near-end and far-end currents i_n and i_f in circuit b at the ends of dx. Since the coefficients are the crosstalk per mile per kilocycle, the near-end coefficient N and the far-end coefficient F may be expressed as follows:

$$N = \text{limit of } \frac{i_n}{I_a} \cdot \frac{10^6}{Kdx} \text{ as } dx \text{ approaches zero.}$$

$$F = \text{limit of } \frac{i_f}{I_a} \cdot \frac{10^6}{Kdx} \text{ as } dx \text{ approaches zero.}$$

where K is the frequency in kilocycles. For circuits of different characteristic impedances Z_a and Z_b the above current ratios should be multiplied by the square root of the ratio of the real parts of Z_b and Z_a. This correction is not included in the expressions for N and F derived below.

Figure 31 indicates the equivalent electromotive forces which, if impressed on the disturbed circuit b, would cause the same direct crosstalk currents as the electric and magnetic fields of the disturbing circuit. The series and shunt electromotive forces V_m and V_e corre-

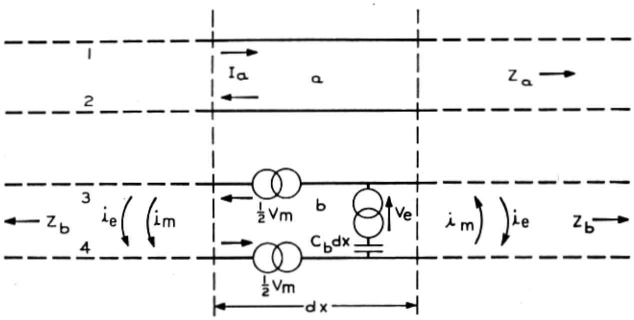

Fig. 31—Equivalent e.m.f.'s in a disturbed circuit.

spond to the magnetic and electric components of the field and cause crosstalk current i_m and i_e. These currents are about equal in magnitude and they add almost directly at the near end of the length dx and subtract almost directly at the far end. The near-end coefficient is, therefore, inherently much greater than the far-end coefficient.

To calculate i_e, the crosstalk current due to the electric field of circuit a, it is necessary to know the shunt voltage V_e. This depends on the charges on the wires of circuit a in the length dx. These charges are due to a voltage V impressed on the left-hand end of circuit a which may be remote from the length dx. Since it is desired to transmit on the metallic circuit a and not on the circuit composed of its wires with ground return, care is taken to "balance" the impressed voltage, i.e., this sending circuit has equal and opposite voltages between its two sides and ground with circuit a disconnected.

The impressed voltage V is propagated to the left-hand end of dx. Letting V_a be the voltage across circuit a at this point, it will be shown that V_a would be balanced except for the effect of interaction crosstalk which is excluded from consideration for the present. Designating the wires of circuit a as 1 and 2, the balanced voltage V_a causes charges Q_1 and Q_2 per unit length on these wires in the length dx. These charges are affected by the presence of other wires in the length dx and they are usually unbalanced. There will be equal and opposite or balanced charges $\pm \frac{Q_1 - Q_2}{2}$ on each wire and unbalanced equal charges $\frac{Q_1 + Q_2}{2}$ on each wire. Since the direct crosstalk is defined as the effect of balanced charges and currents, only the balanced charges should be considered in computing V_e. Letting $Q_a = \frac{Q_1 - Q_2}{2}$ or the balanced charge on wire one per unit length, then:

$$Q_a = V_a C_a = I_a Z_a C_a,$$

where C_a is equal to the "transmission capacitance" per mile, i.e., the capacitance used in calculating α_a the attenuation constant and Z_a the characteristic impedance of circuit a.

The above expression for Q_a includes the reaction of charges in the disturbed circuit. This reaction should not theoretically, be included at this time, since, for convenience in calculation, the disturbed circuit is assumed to have the impressed voltages V_m and V_e but no crosstalk currents or charges as yet. The effect on Q_a of charges in the disturbed circuit, is, however, usually small compared with the effect of charges in various tertiary circuits.

Designating the conductors of circuit b as 3 and 4, V_e is the difference of the potentials of the electric field at 3 and 4 caused by the balanced charges per unit length on 1 and 2. Therefore:

$$V_e = V_3 - V_4 = (Q_a p_{13} - Q_a p_{23}) - (Q_a p_{14} - Q_a p_{24}),$$

where p_{13}, etc., are the potential coefficients.

For c.g.s. elst. units, $p_{13} = 2 \log \frac{s_{13}}{r_{13}}$ where s_{13} and r_{13} are the distances indicated by Fig. 32. Therefore:

$$V_e = V_a C_a (p_{13} - p_{23} - p_{14} + p_{24}) = V_a C_a p_{ab}.$$

The capacitance C_a may be obtained from measurements on a short length of a multi-wire line. Its value is, however, only a few per cent greater than C_a' the value for a single pair line (without capacitance at the insulators). For a single pair having like wires in a horizontal

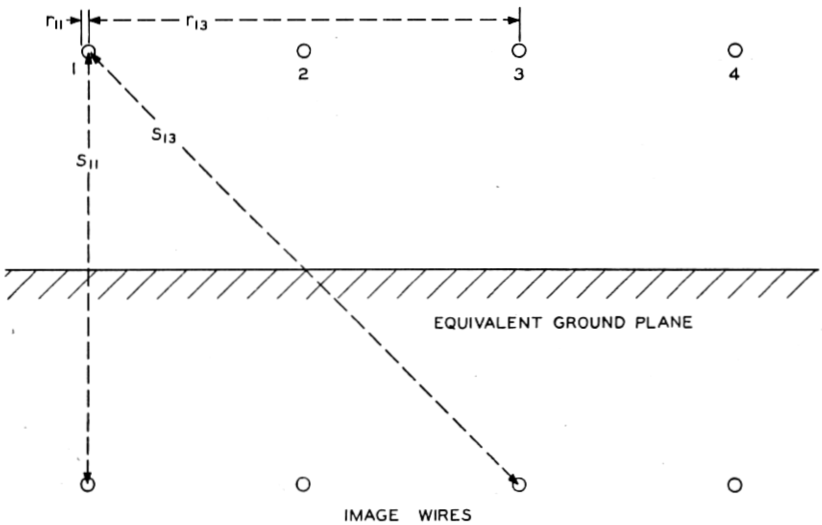

Fig. 32—Distances used in computing potential coefficients.

plane, C_a' is readily calculated as follows:

$$C_a' = \frac{1}{2(p_{11} - p_{12})},$$

where p_{11} in c.g.s. elst. units is:

$$2 \log_\epsilon \frac{s_{11}}{r_{11}} \cdot$$

The distances s_{11} and r_{11} are indicated on Fig. 32.

The expression for V_e may be written:

$$V_e = V_a C_a p_{ab} = V_a C_a' p_{ab} \frac{C_a}{C_a'} = V_a T_{ab} \frac{C_a}{C_a'} \cdot$$

The coefficient T_{ab} is called the "voltage transfer coefficient." It is readily computed since it is a function of potential coefficients and it is independent of the system of units used in computation. Since C_a is about equal to C_a', T_{ab} is about equal to the ratio of V_e to V_a.

The shunt voltage V_e drives a current through the shunt admittance of circuit b in the infinitesimal length dx of Fig. 31. This shunt admittance is $(G_b + j\omega C_b)dx$ which is very nearly equal to $j\omega C_b dx$ where C_b is the transmission capacitance per mile of circuit b and

$\omega = 2\pi f$ where f is the frequency in cycles per second. This current divides equally between the two ends of circuit b. The near-end current is:

$$i_e = -\frac{1}{2}\frac{V_e}{\dfrac{1}{j\omega C_b dx} + \dfrac{Z_b}{2}}.$$

The near-end direct crosstalk coefficient due to the electric field of circuit a may be called N_e and is the limiting value of the following expression as dx approaches zero:

$$\frac{i_e}{I_a}\cdot\frac{10^6}{Kdx} = -\frac{1}{2KI_a}\cdot\frac{V_e 10^6}{\dfrac{1}{j\omega C_b} + \dfrac{Z_b}{2}dx},$$

where K is the frequency in kilocycles. The near-end direct crosstalk coefficient for the electric field is, therefore:

$$(1)\quad N_e = -\frac{V_e 10^6 j\omega C_b}{2KI_a} = -\frac{Z_a T_{ab} j\omega C_b 10^6}{2K}\cdot\frac{C_a}{C_a'}$$

$$= -j\pi Z_a T_{ab} C_b 10^9 \frac{C_a}{C_a'}.$$

The far-end current due to V_e of Fig. 31 is $-i_e$ and, therefore, the far-end coefficient due to the electric field is $-N_e$.

The near-end and far-end crosstalk currents of Fig. 31 due to the magnetic field are alike and are designated i_m which may be calculated as follows:

$$i_m = \frac{V_m}{2Z_b} = -\frac{I_a j\omega M_{ab} dx}{2Z_b}.$$

The near-end or far-end crosstalk coefficients for the magnetic field may be called N_m and F_m. They are alike and equal to the limit of:

$$\frac{i_m}{I_a}\cdot\frac{10^6}{Kdx} \text{ as } dx \text{ approaches zero.}$$

Therefore:

$$N_m = F_m = -\frac{j\omega M_{ab}}{2Z_b K}10^6 = -\frac{j\pi M_{ab}}{Z_b}10^9.$$

In the above, M_{ab} is the mutual inductance per unit length between circuits a and b. It is calculated in the same manner as p_{ab} used in computing V_e. These methods of computing V_e and V_m from the distances r_{13}, r_{14}, s_{13}, etc., of Fig. 32 are not precise but are sufficiently accurate for open-wire circuits since the diameters of the wires are small compared with their interaxial distances. The "image" wires of Fig. 32 should, theoretically, be located farther below the equivalent ground plane for calculations of mutual inductance. This alters s_{14}, etc. Since the distances between wires are small compared to those between wires and images, the values of s are all about equal and have practically no effect on the value of p_{ab}.

Therefore, M_{ab} in c.g.s. elmg. units may be assumed numerically equal to p_{ab} or:

$$M_{ab} = p_{ab} = \frac{T_{ab}}{C_a'}.$$

In c.g.s. elst. units $C_a' = \dfrac{1}{2(p_{11} - p_{12})}$ which is also the expression for $1/L_a'$ in c.g.s. elmg. units where L_a' is the *external* inductance of circuit a, i.e., the inductance due to the magnetic field external to the wires of circuit a. Therefore:

$$M_{ab} = T_{ab}L_a',$$

where M_{ab} and L_a' may be expressed in any system of units.

The near-end or far-end direct coefficient for the magnetic field may, therefore, be written:

$$N_m = F_m = -\frac{j\pi T_{ab}L_a'}{Z_b}10^9. \tag{2}$$

The above expression is almost equal to N_e, the near-end coefficient for the electric field.

It may be written:

$$N_m = N_e\left[\frac{j\omega L_a' j\omega C_a'}{Z_a j\omega C_a Z_b j\omega C_b}\right].$$

Now $Z_a j\omega C_a$ is very nearly equal to $Z_a(G_a + j\omega C_a)$ which is γ_a. Likewise $Z_b j\omega C_b$ is very nearly equal to γ_b. If the circuits had no resistance or leakance the propagation constant would be $\gamma_0 = j\omega\sqrt{L_a'C_a'}$ or $j\omega/v$ where v is the speed of light in miles per second. Therefore:

$$N_m = \frac{N_e\gamma_0^2}{\gamma_a\gamma_b}\text{ very nearly.}$$

The total direct crosstalk coefficients are:

$$N_d = N_m + N_e = N_e\left(1 + \frac{\gamma_0^2}{\gamma_a\gamma_b}\right) = 2N_e\text{ approx.} \tag{3}$$

At carrier frequencies the ratio of γ_0 to γ_a (or γ_b) is about equal to the ratio of the actual speed of propagation to the speed of light, i.e., 180,000 to 186,000 or about .97. Therefore $\left(1 + \frac{\gamma_0^2}{\gamma_a \gamma_b}\right)$ is about 1.94.

$$(4) \quad F_d = N_m - N_e = N_e \left(\frac{\gamma_0^2}{\gamma_a \gamma_b} - 1 \right)$$

$$= N_e \left(-.06 + j \frac{90}{\pi} \frac{\alpha_a + \alpha_b}{K} \right) \text{ approx.}$$

The attenuation of the disturbing and disturbed circuits may not be neglected in evaluating the expression $\left(\frac{\gamma_0^2}{\gamma_a \gamma_b} - 1 \right)$.

The expression given for N_e in equation (1) above may be written:

$$N_e = -\frac{\gamma_a T_{ab} 10^6}{2K} \frac{C_a}{C_a'} \frac{C_b}{C_a}.$$

This assumes $Z_a j\omega C_a$ equals γ_a. At carrier frequencies γ_a is about equal to $j\beta_a$ which is about $j\pi K/90$ since the speed of propagation is about 180,000 miles per second. The expression for N_e may, therefore, be written in the following simple approximate form:

$$N_e = -j \frac{\pi T_{ab} 10^6}{180} \frac{C_a}{C_a'} \cdot \frac{C_b}{C_a}.$$

The ratio of C_a to C_a' does not ordinarily exceed 1.02. For like circuits, therefore:

$$N_e = \frac{-j\pi T_{ab} 10^6}{180} \text{ approx.}$$

On Fig. 33 the magnitudes of N_d and F_d are plotted against frequency for 8-inch spaced conductors .128-inch in diameter. Both F_d and N_d are divided by T_{ab} to make the curves applicable to any circuit combination. These curves show that N_d is practically independent of frequency (above a few hundred cycles) but F_d decreases rapidly with frequency for several thousand cycles.

Indirect Crosstalk Coefficients

Expressions for the indirect crosstalk coefficients used in computing the indirect component of transverse crosstalk coupling will now be derived. The derivation first covers the case of a single representative tertiary circuit. Fig. 34 shows a thin transverse slice of the parallel of the three circuits, the thickness of the slice being the infinitesimal length *dx*. The only tertiary circuit to be considered for the present is the metallic circuit composed of wires 5 and 6 and designated as *c*. There are other possible tertiary circuits in the system of 6 wires, for example, the phantom circuit composed of wires 1 and 2 as one side and 5 and 6 as the other side. The method of estimating the total effect of all possible tertiary circuits will be discussed later.

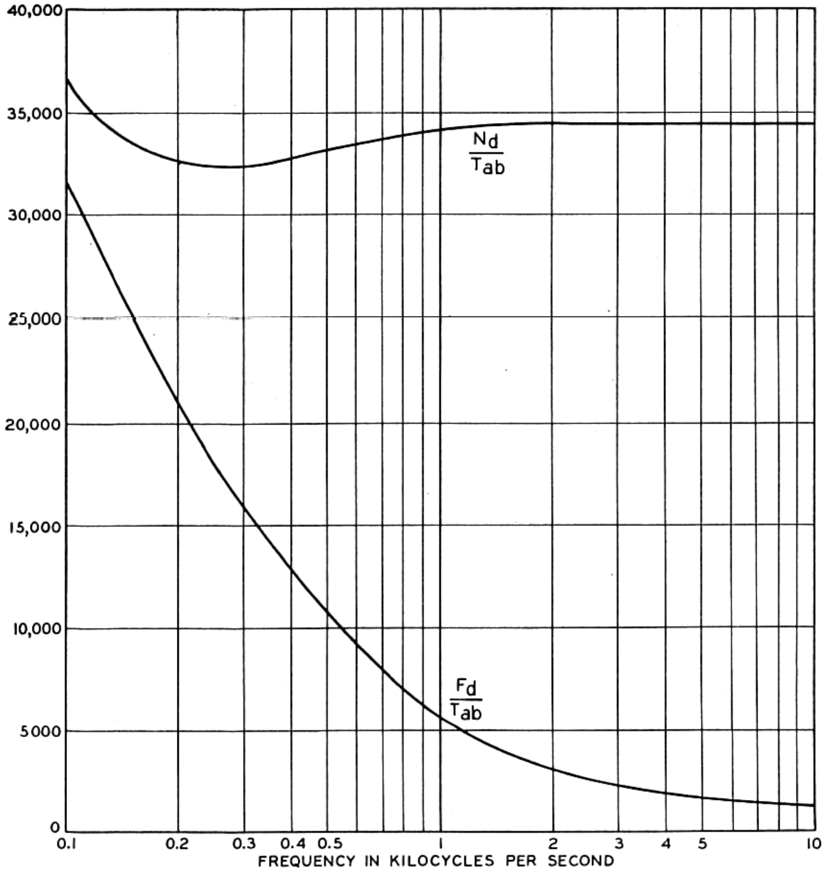

Fig. 33—Variation of direct crosstalk coefficients with frequency.

The immediate problem is to compute the crosstalk currents in circuit *b* at the ends of the length *dx* due to currents and charges in circuit *c* in this length and caused by transmission over circuit *a* through *dx*.

The crosstalk currents in circuit *b* due to currents and charges in circuit *a* were computed by determining the equivalent series and shunt e.m.f.'s in circuit *b*. The effect of currents and charges in circuit *c* on crosstalk currents in circuit *b* may be computed in a similar manner. The series e.m.f. in circuit *b* proportional to the current in circuit c will, however, be negligible compared with the series e.m.f. proportional to current in circuit a. This is evident since the current in circuit c is a crosstalk current which approaches zero as dx approaches zero while the current in circuit a does not vary with dx.

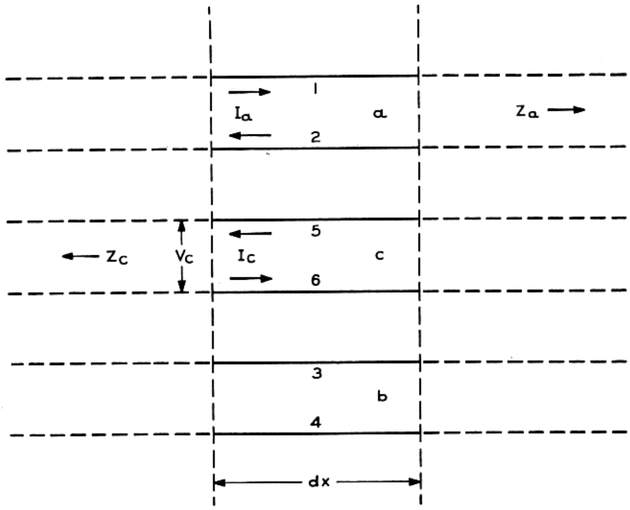

Fig. 34—Schematic used in deriving formulas for indirect crosstalk coefficients.

The shunt e.m.f. in circuit b dependent on the charges of circuit c is not, however, negligible compared with the shunt e.m.f. in circuit b due to charges in circuit a since the charges in both a and c approach zero as dx decreases. In other words the magnetic field of circuit c may be neglected but the electric field must be considered. (Both fields must be considered in computing interaction crosstalk.)

To determine the equivalent shunt e.m.f. in circuit b which depends upon the electric field of circuit c the voltage between the wires of circuit c must be determined. If circuit c did not exist, the electric field of circuit a would cause a difference of potential between the points actually occupied by wires 5 and 6 at the left-hand end of dx in Fig. 34. This difference of potential would be:

$$V_{ac} = V_a C_a' p_{ac} = V_a T_{ac}.$$

With circuit c present, this difference of potential is changed to V_c, the actual voltage across circuit c. The voltage could not change from V_{ac} to V_c without charges on circuit c and the charge per wire per unit length is proportional to the change in voltage from V_{ac} to V_c which may be designated U_c. The equivalent shunt e.m.f. in circuit b due to the presence of charges in circuit c is, therefore, proportional to U_c.

By definition:

$$V_{ac} + U_c = V_c \quad \text{or} \quad U_c = V_c - V_{ac}.$$

Since the crosstalk current in circuit c approaches zero as dx approaches zero, V_c must also approach zero and U_c approaches $-V_{ac}$. The shunt e.m.f. in circuit b due to charges on circuit a was computed as:

$$V_e = V_a T_{ab} \frac{C_a}{C_a'}.$$

To allow for the *electric* field of circuit c, V_e must be augmented by:

$$V_e' = U_c T_{cb} \frac{C_c}{C_c'} = -V_{ac} T_{cb} \frac{C_c}{C_c'} = -V_a T_{ac} T_{cb} \frac{C_c}{C_c'}.$$

Since the part of the direct near-end crosstalk coefficient resulting from V_e was found to be $N_e = -j\pi Z_a T_{ab} C_b 10^9 \dfrac{C_a}{C_a'}$, by proportion the indirect near-end coefficient resulting from V_e' will be:

(5) $$N_i = j\pi Z_a T_{ac} T_{cb} C_b 10^9 \frac{C_c}{C_c'} = \frac{j\pi T_{ac} T_{cb} 10^6}{180} \text{ approx.}$$

Since the far-end crosstalk current resulting from a shunt voltage in circuit b is opposite in sign to the near-end current, the indirect far-end coefficient will be:

(6) $$F_i = -N_i.$$

Total Crosstalk Coefficients

The total near-end and far-end crosstalk coefficients used in computing transverse crosstalk coupling will be the sum of the direct and indirect coefficients or:

(7) $$N = N_d + N_i.$$

(8) $$F = F_d + F_i = F_d - N_i.$$

The expressions for F_i and N_i are about independent of frequency in the carrier-frequency range because Z_a does not depend much on frequency above a few thousand cycles, C_b is about independent of frequency and $T_{ac}T_{cb}$ depends only on the cross-sectional dimensions of the wire configuration.

Since, as indicated by Fig. 33, N_d is usually about independent of frequency and since $N = N_d + N_i$ is largely determined by N_d, the near-end coefficient N is about independent of frequency above a few hundred cycles. The far-end coefficient F is about independent of frequency above a few thousand cycles where it is largely determined by F_i.

The preceding discussion of indirect crosstalk coefficients covered only the effect of charges in the single metallic tertiary circuit c of Fig. 34. The indirect coefficient in a practical case may be estimated with fair accuracy by considering all the more important tertiary circuits in a similar manner. It was shown that the final voltage of tertiary circuit c was zero. Similarly, the final voltage of each tertiary circuit is zero. This includes any tertiary circuits involving the two wires of the disturbing circuit in multiple. The average voltage of the two wires of the disturbing circuit is zero and the voltage across the disturbing circuit is balanced. As previously stated, this voltage does not become unbalanced as a result of transverse crosstalk in any infinitesimal length but it may become unbalanced due to interaction crosstalk.

The charges per unit length on the various tertiary circuits are the same as those which would be caused by impressing a system of voltages equal and opposite to those induced by the balanced charges per unit length which would be on the two wires of the disturbing circuit if this circuit were the only pair on the line. Assuming such a system of impressed voltages, it is not practicable to accurately compute the charges in any tertiary circuit since this depends on the voltages impressed on all the tertiary circuits and the couplings between the various tertiary circuits. Advantage may be taken, however, of the fact that the charge on a tertiary circuit will depend mostly on the voltage impressed on that circuit provided it is not heavily coupled with other circuits.

It is possible to divide the various voltages impressed on the tertiary circuits into components such that (1) equal voltages are impressed on wires of a "ghost" circuit composed of all the wires on the line with ground return, (2) balanced voltages are impressed on each pair used for transmission purposes (except the disturbed and disturbing circuits) and (3) balanced voltages are impressed on each possible phantom of two pairs used for transmission purposes.

Such a system of impressed voltages and tertiary circuits is convenient for computation since the charge on any tertiary circuit largely depends on the voltage impressed on that circuit. If accurate calculations of the charges were practicable, a simpler system of tertiary circuits could be used to obtain the same final result, i.e., single-wire tertiary circuits with ground return could be used. Computation with such tertiary circuits is impracticable because of the large coupling between them.

In the elaborate system of tertiary circuits described above, the ghost circuit may be neglected. The voltage impressed across this circuit is the average of all the voltages impressed on the various wires. These voltages may be plus or minus and the average tends to be small. Also, the charge per pair per impressed volt is usually much less for the ghost circuit than for a phantom circuit due to the relatively small capacitance between a pair and ground as compared with that between two pairs.

The pairs used for transmission purposes may usually be disregarded, also, since their coupling with the disturbing and disturbed circuits is much smaller than that of the phantom tertiary circuits.

The practical method of computing the indirect crosstalk coefficient is, therefore, to consider as tertiary circuits a considerable number of phantoms composed of pairs used for transmission purposes including the disturbed and disturbing pairs. In calculating the charge in any tertiary circuit, the voltages impressed on other tertiary circuits are disregarded.

In calculating the effect of a single tertiary circuit c, the expression for the indirect coefficient contained the factor $T_{ac}T_{cb}$. To estimate the effect of all the tertiary circuits, this factor should be replaced by:

$$\frac{2}{m}\sum T_{ap}T_{pb}.$$

This expression assumes that there are n pairs on the line and that $m - 2$ of these pairs are close enough to the disturbing and disturbed pairs to appreciably affect the indirect crosstalk between them. The subscript p indicates any phantom of the m pairs including the disturbed and disturbing pairs. The summation is for all possible phantoms each consisting of two of the m pairs. If the voltages induced by the balanced charges Q_a' of pair a are V_r and V_s for the two sides of a phantom, the balanced voltage assumed to be impressed across the phantom is $\frac{2}{m}(V_s - V_r)$. Other parts of V_r and V_s are used in the "ghost" voltage and in balanced voltages across other phantoms.

T_{ap} and T_{pb} are voltage transfer coefficients relating balanced impressed voltage on the disturbing circuit to induced voltage on the disturbed circuit. T_{ap} involves C_a' the transmission capacitance of circuit a on a single pair line. T_{pb} involves the transmission capacitance of a particular phantom on a line having only that phantom present. This capacitance is the ratio of balanced charge (on each side of the phantom) to the balanced impressed voltage. The phantom capacitance may be readily estimated from the potential coefficients. For example, if the phantom involves pairs 1–2 and 5–6 the phantom capacitance is very nearly:

$$C_p = \frac{4}{p_{11} + p_{22} + p_{55} + p_{66} + 2p_{12} + 2p_{56} - 2p_{15} - 2p_{25} - 2p_{16} - 2p_{26}}.$$

If the disturbing circuit is pair 1–2 and the disturbed circuit is pair 3–4:

$$T_{ap} = \frac{p_{15} + p_{16} - p_{25} - p_{26}}{2(p_{11} - p_{12})}$$

$$T_{pb} = \frac{C_p}{2}(p_{13} + p_{23} + p_{45} + p_{46} - p_{14} - p_{24} - p_{35} - p_{36}).$$

These computations of indirect coefficients are necessarily laborious. They can be simplified to some extent by ignoring phantoms for which either T_{ap} or T_{pb} is zero or small. For example, the voltage transfer coefficient is zero for pair 1–2 to such phantoms as 1–2 and 11–12, 11–12 and 21–22, etc.

In the following table are given comparisons of far-end crosstalk coefficients as measured in a 40-wire line and as computed by the methods discussed above. The spacing of the various wires and crossarms is indicated by Fig. 27A. The measured values are for 40 wires and the computed values are for 10, 20 and 30 wires. It will be seen that a considerable number of wires must be taken into account in the computations in order to obtain a fair check with the coefficient measured for a heavy line.

Far-End Crosstalk Per Mile Per Kilocycle

	Combination		
	1–2 to 3–4	1–2 to 11–12	1–2 to 9–10
Computed for 10 wires	45	28	
Computed for 20 wires	63	47	11
Computed for 30 wires	69	58	22
Measured for 40 wires	74	70	21

Interaction Crosstalk Coefficient

It was assumed in the discussion of crosstalk coefficients that the "interaction crosstalk coefficient" $N_{ac}N_{cb}10^{-6}$ was nearly equal to $-2F_i\gamma_c/K$. This relation is deduced below, for a representative tertiary circuit c, from the expressions for F_i and N_d given by equations (3) and (6) above. N_{ac} may be obtained by using the expressions for N_e and N_d given by equations (1) and (3) above. In these equations, subscript c should be substituted for subscript b. The expression for N_{ac} becomes:

$$N_{ac} = -j\pi Z_a T_{ac} C_c 10^9 \frac{C_a}{C_a'}\left[1 + \frac{\gamma_0^2}{\gamma_a\gamma_c}\right].$$

Deriving a similar approximate expression for N_{cb}:

$$N_{ac}N_{cb}10^{-6} = -\frac{F_i}{2K}\gamma_c\frac{C_a}{C_a'}\left[1 + \frac{\gamma_0^2}{\gamma_a\gamma_c}\right]\left[1 + \frac{\gamma_0^2}{\gamma_c\gamma_b}\right].$$

This assumes $Z_c j\omega C_c = Z_c(G_c + j\pi C_c)$ which is γ_c.

Crosstalk measurements indicate that the ratio of γ_0 to γ_a or γ_b or γ_c is about .97 at carrier frequencies and C_a/C_a' about 1.02. Therefore:

$$N_{ac}N_{cb}10^{-6} = -\frac{F_i\gamma_c}{2K}1.02(1.94)^2 = -2F_i\gamma_c/K \text{ approx.}$$

The above discussion covers the case of a single tertiary circuit c. In the practical case the known crosstalk coefficient F_i includes the effect of a large number of tertiary circuits which have various values of γ_c. Obtaining the interaction crosstalk coefficient from the expression $-2F_i\gamma_c/K$ involves assuming a representative value of γ_c. At carrier frequencies γ_c is about equal to $j\beta_c$ which for all important tertiary circuits is in the neighborhood of $j\pi K/90$.

A known value of the crosstalk coefficient F_i expresses the effects of the electric fields of many tertiary circuits in one infinitesimal slice and includes the alteration of the field of any one tertiary circuit due to the presence of the others. The interaction crosstalk coefficient involves consideration of both electric and magnetic fields. In any one slice, the electric field of a tertiary circuit determines its crosstalk into the disturbed circuit but the current in the tertiary circuit at the end of the slice is determined by both the electric and magnetic fields of the disturbing circuit. The tertiary circuit current is transmitted into another slice and sets up electric and magnetic fields which both contribute to the crosstalk current in the disturbed circuit.

In the case of a single tertiary circuit, the electric and magnetic effects expressed by the interaction coefficient are simply related and the interaction coefficient is simply related to F_i. With many tertiary circuits the relation between the electric and magnetic fields of any one tertiary circuit is altered by the presence of the others and the relative importance of electric and magnetic fields in the interaction effect varies among the tertiary circuits (except for the ideal case of "non-dissipative" circuits). In a practical case, therefore, the interaction coefficient is only approximately proportional to F_i. Measurements of the interaction coefficients by indirect methods have, however, indicated that the approximation is satisfactory for purposes of practical transposition design.