The E6 Negative Impedance Repeater

By A. L. BONNER, J. L. GARRISON and W. J. KOPP

(Manuscript received April 14, 1960)

The E6 repeater is a low-cost, transistorized, voice-frequency, two-way repeater for use in the exchange area plant, including use in trunks between local offices and between local and toll offices. It permits both types of trunks to be operated at low net loss, while meeting the impedance requirements placed on trunks between local and toll offices. These features are possible because of a unique method of repeater design, in which the impedance-matching function and the gain function are allocated to separate portions of the repeater and are separately adjustable.

The impedance-matching function is accomplished by means of a line building-out network, which permits any cable pair associated with the repeater to be built out and made to match the impedance of the toll office. The gain unit presents the same image impedance, which is essentially independent of gain adjustment. Changes in gain are controlled by two resistive, adjustable networks — one in the series-connected negative impedance converter, and the other in the shunt-connected converter. In previous designs of negative impedance repeaters, complex impedance networks were required for adjustment of both gain and impedance match.

The E6 repeater also incorporates novel mechanical features, with transistorization resulting in low power drain and less heat dissipation in the telephone central office. Power is derived from the existing 48-volt central office battery.

Table of Contents

I. INTRODUCTION

1.1 *The E23 Negative Impedance Electron Tube Repeater*

One of the methods used to reduce the transmission losses in exchange area cable circuits employs E23* negative impedance repeaters. These repeaters,[1] first installed in 1954, represented a considerable advance over the series-type negative impedance repeaters[2] then in use. The improvement was accomplished by the addition of a shunt-type converter to the existing series-connected negative impedance element. Impedance discontinuities introduced into otherwise uniform loaded circuits and the resulting undesirable reflection effects were thereby avoided. The effectiveness of this technique is attested by the fact that over one million E repeater units have been installed in the Bell System since 1954.

Installation of the E23 repeater is relatively complex, since the negative impedance of the repeater must be closely matched to the cable impedance in order to give the desired gain and return loss characteristic. The matching is accomplished by strapping components of complex networks included in the repeater. The required strapping varies with (a) the amount of gain required, (b) the cable facility with which the repeater is associated and (c) the length of cable end-section to which the repeater is connected. Network connections for a particular condition are specified by extensive strapping charts. This is a time-consuming operation on the initial installation and on subsequent changes, either for gain or end-section adjustment.

A basic limitation on the application of the E23 repeater to telephone offices is that it is an electron tube device requiring both a 130-volt DC

* An E23 repeater is made up of an E2 series-type repeater and an E3 shunt-type repeater connected in a bridged-T arrangement.

plate supply and a 48-volt DC filament source. The 48-volt supply is almost universally available in telephone offices, but the necessity for a 130-volt plate supply has limited the use of this repeater to offices where such voltage supplies exist, or where the cost of installing a reliable 130-volt source could be justified.

1.2 *The E6 Negative Impedance Transistorized Repeater*

In reviewing the disadvantages of the E23 repeater it was apparent that by use of transistors in the gain portion of the repeater the need for a 130-volt supply could be eliminated, operating costs be drastically reduced and the annoying problem of heat dissipation in large installations be largely removed. Design studies also indicated that definite advantages would result if it were possible to associate with the repeater a wideband matching network that would match the complex impedance of the associated cable facility to the image impedance of the converter section of the repeater. By thus simplifying the impedance matching requirements on the converter, it appeared possible to develop a negative impedance repeater requiring only resistive networks for gain and return loss adjustment. Thus the concept of the E6 repeater was evolved.

II. SYSTEMS APPLICATIONS

2.1 *General Description*

The application of negative impedance repeaters to the exchange area plant is dependent on the type of service performed by the trunk facilities with which they are to be associated. If a repeater introduces an impedance discontinuity into an otherwise uniform loaded circuit, a substantial amount of energy will be returned to the sending end as "echo." The quality of transmission is related to the magnitude of this echo and to the delay introduced by the transmission facility before the echo reaches the talker's ear. As a result, when the delay is small, as in the case of an interoffice trunk for short-haul service, a relatively large amount of echo can be permitted. On toll-connecting trunks, however, the delay may be large and only a small amount of echo can be allowed. Since the gain obtainable from a series-type negative impedance repeater is dependent on the amount of negative impedance inserted in series with the trunk, it follows that the resulting echo is proportional to the gain of the repeater. Application of the series-type negative repeater is, therefore, limited, except at low gains, to interoffice trunk use. The E23 repeater, because of its better return loss, can be used with correspondingly higher gains. The E6 repeater is even more flexible in specific applications inasmuch as its maximum gain is largely determined by the structural return losses of the connecting circuits rather than by repeater characteristics. Later in the discussion, the performance of the E23 and E6 repeaters will be compared in detail.

2.2 *Principles of Operation — E23 and E6 Repeaters*

The block diagram of Fig. 1 illustrates the principles of the E23 repeater. The single block labelled E23 CONVERTER contains the negative impedance controlling features of the repeater. In the diagram, the E23 repeater is shown at the local office end of a toll-connecting trunk circuit. Inserted between the line and the toll office is a standard impedance compensator used to improve the match between the office and the loaded cable facility. Branching out from the repeater, but included as part of the series and shunt converters comprising it, are two variable, complex impedance networks. These networks are designed so that the impedance of the converter over the voice-frequency band, for a particular gain adjustment, closely approximates the impedance of the connected cable facility. Element values are obtained by strapping of passive elements selected from the 51 elements available within the repeater.

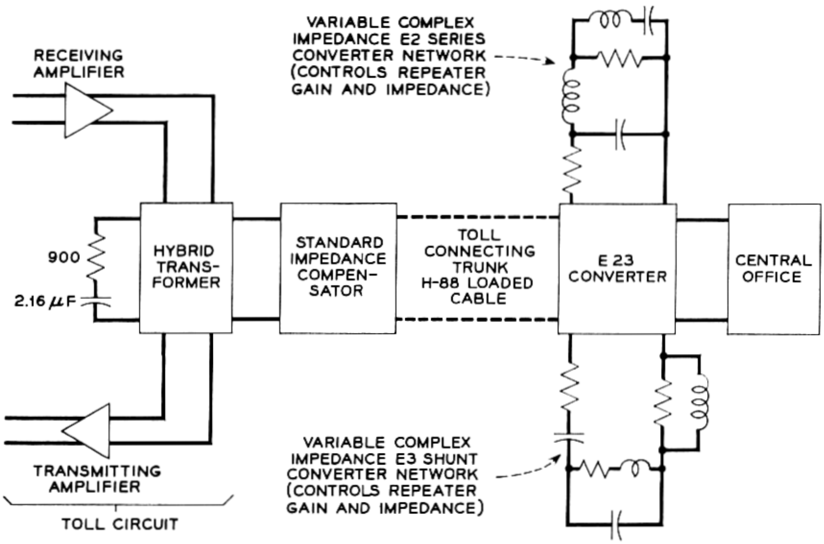


Fig. 1 — E23 repeater installation; terminal use — toll-connecting trunks.

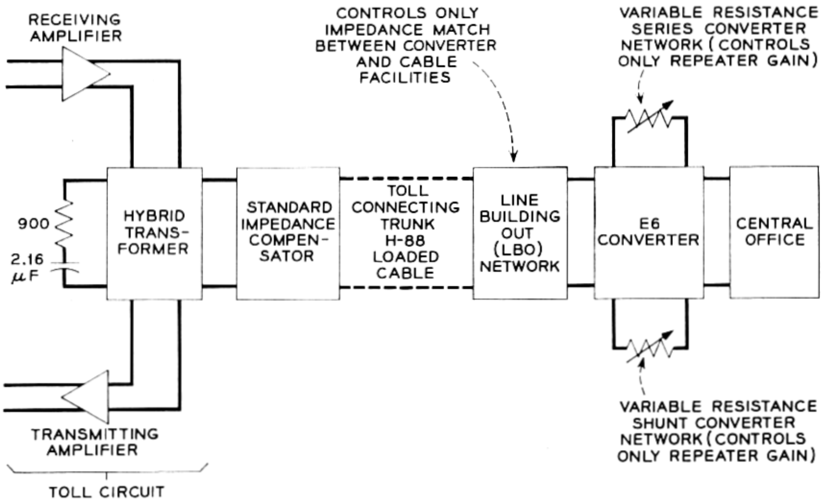


Fig. 2 — E6 repeater installation; terminal use — toll-connecting trunks.

For comparison purposes, the application of the E6 repeater to the same situation is shown in Fig. 2. It should be re-emphasized that the configuration of the two converters and the philosophy underlying the two designs are not the same. The complex impedance networks used in the E23 repeater have been reduced in the E6 repeater to simple adjustable resistances, and a wideband matching or line building-out network (LBO) has been added to match the impedance of the converter to that of the trunk. Inasmuch as the E6 repeater is situated at the local office end, an impedance compensator is still required at the toll office end to improve the match to the office impedance. The adjustment of the line building-out network of the E6 repeater is a simple operation involving the loosening or tightening of a number of screw connections. The gain of the repeater is similarly adjusted. This compares to strapping and soldering a number of terminals on the E23 repeater.

It should be pointed out, however, that the E6 repeater, although more efficient and easier to adjust, does not provide the wide flexibility of use inherent in the E23 repeater. A separate wideband matching network is required for each type of loaded facility to which it is to be connected, whereas the E23 repeater can approximate the impedance of a wide variety of circuits. Individual networks can, however, be provided containing only those elements required for a particular line facility. Fortunately, more than 85 per cent of existing exchange area circuits can be handled by one network, namely, that designed for H-88 loaded cable.

Although the physical arrangement of the E6 repeater will be considered in detail in a later section, photographs of the E23 and E6 repeaters are shown in Fig. 3 for immediate comparison. The E23, as shown at the right, is mounted in two separate die-cast housings, one for the E2 series-connected converter and its associated network, and the other for the E3 shunt-connected converter and its associated network. Terminals for network strappings to control both gain and impedance are clearly indicated on the E2 repeater. Similar terminal arrangements exist on the E3.

The E6 repeater, arranged with two LBO networks for insertion at the junction of two sections of cable (intermediate use) is shown at the left of the photograph. The LBO networks are mounted in the top of an extruded aluminum H-beam to which a front cover is attached. The E6 converter is mounted at the bottom of the H-section extrusion. The

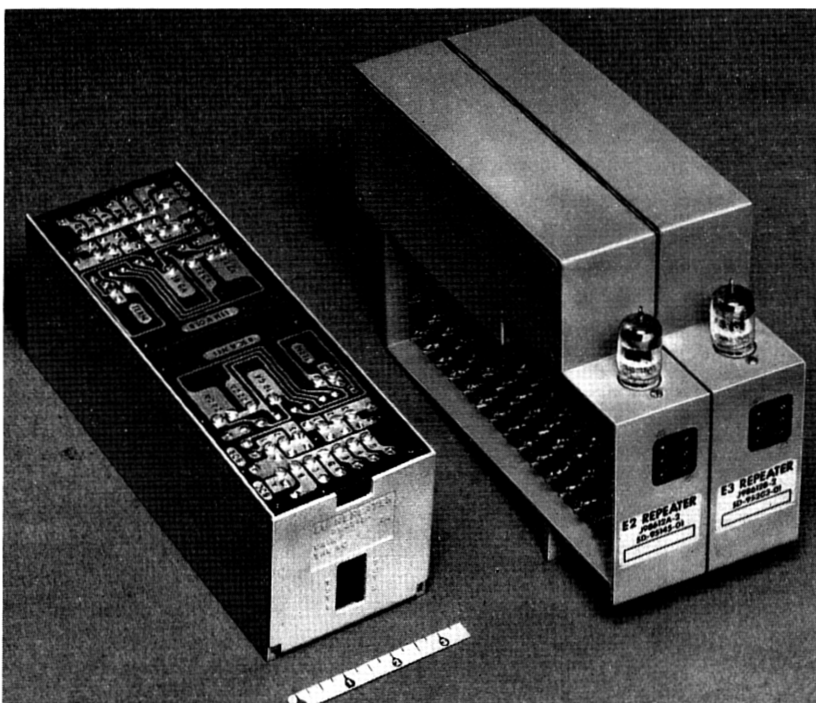


Fig. 3 — E6 repeater (left); E23 repeater (right).

screw-type terminals used on the LBO networks for impedance adjustment are indicated. Similar screw-type adjustment features are provided on the converter for gain adjustment.

III. DESIGN OBJECTIVES

In establishing a plan for achieving the important improvements and economies believed to be attainable by use of transistorized negative impedance converters in association with wideband matching networks, it was realized that the new negative impedance repeater would be competing with an established product that was easy to manufacture and free from early production difficulties. Considerations relating to cost and to ease of manufacture were therefore of primary importance. From a production point of view it was also deemed essential that the latest manufacturing techniques be applied and that full consideration be given to the possibility of automated processes. The availability of low-cost, long-lived transistors was essential to the realization of the desired economies. Since transistors are inherently low-power devices, protection had to be provided against the effects of (a) foreign potentials in the telephone plant due to lightning and induced surges from power circuits and (b) high potentials due to signaling. Such surges may occur from either tip and ring wires to ground or between tip to ring wires. Surge peaks greater than 600 volts are normally limited by carbon-electrode discharge gaps, so that any repeater must be designed to tolerate voltages up to this magnitude.

Electrical requirements were established to meet the projected needs of the Bell System exchange area plant. In recent years there has been the objective of reducing the losses in this area of the plant in order to improve subscriber-to-subscriber transmission. Today's objective for direct trunks is 4 db average, while 2 db to 3 db maximum is required for tandem or toll-connecting trunks. Broadly interpreted, two objectives were outlined for the repeater:

i. To provide a transistorized voice-frequency negative impedance repeater having series and shunt converters (in association with wideband matching networks) so arranged as to reduce circuit net losses in discrete adjustable steps by means of variable resistance networks and to present an image impedance equal to 900 ohms in series with 2.16 microfarads; i.e., 900 + 2.16 microfarads.

ii. To provide a wideband impedance matching network (line building-out network) capable of matching variable end section loaded cable facilities of different gages to the impedance presented by the converter section of the repeater under conditions of terminal or intermediate use.

IV. THEORY

4.1 *General*

Negative impedance repeaters such as the E23 and E6 are based on a bridged-T network configuration with mutual coupling. The standard bridged-T configuration for the case of positive impedance elements is shown in Fig. 4(a). The image impedance and insertion loss of the network are seen to be simply related to the impedances Z_A and Z_B of the series and shunt elements comprising it. Similar relations, indicated in Fig. 4(b), exist for the case of negative impedance elements, and the defining equations for image impedance and insertion gain are identical in form to the previous case except that substitution of negative impedance elements results in a gain instead of a loss. This is more apparent if the negative sign of the series arm is written as

$$Z_A \overline{\vert 180^\circ}$$

and the shunt arm as

$$Z_B \underline{\vert 180^\circ}.$$

4.2 *The E23 Repeater*

If a negative impedance repeater were to be inserted into a distortion-

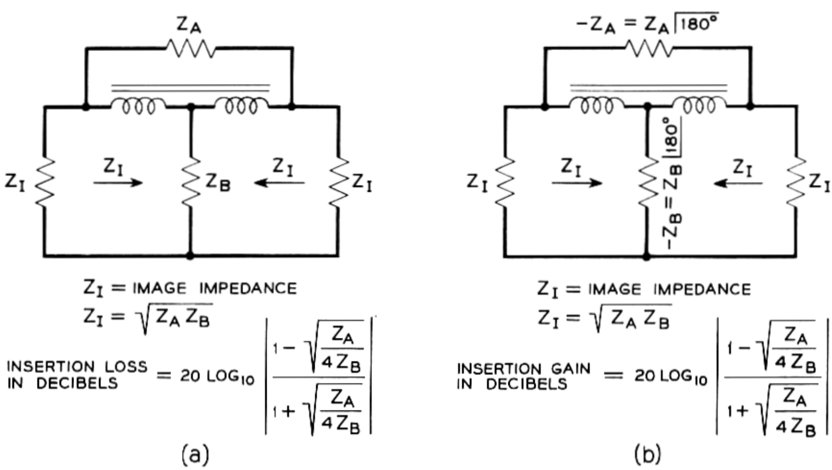


Fig. 4 — (a) Bridged-T positive impedance network; (b) bridged-T negative impedance network.

less transmission line, without introducing impedance discontinuities, the image impedance of the repeater would have to match the characteristic impedance of the line. In the E23 repeater this impedance match was established by properly relating the series and shunt elements of the equivalent bridged-T structure to the characteristic impedance of the line by a proportionality constant N. Assuming that the negative series impedance Z_A of Fig. 4(b) is equated to $NZ_0 \overline{|180°}$, and that the negative shunt impedance Z_B is set equal to $Z_0/N \underline{|180°}$, the image impedance of the repeater, Z_I, may be written

$$Z_I = \sqrt{Z_A Z_B} = Z_0 ,$$

where $Z_0 =$ the characteristic impedance of the line.

Similarly, if the above expressions for Z_A and Z_B are substituted into the equation for insertion gain given in Fig. 4(b), the insertion gain may be expressed as

$$\text{insertion gain (in db)} = 20 \log_{10} \left| \frac{1 + N/2}{1 - N/2} \right| .$$

The above expression for insertion gain may be used to determine stability limits for such repeaters. If $N = 2$, the insertion gain is infinite and the repeater will sing. The inherent limitations on available gain from negative impedance repeaters was recognized at an early stage. In 1919, while presenting an historic account of the research and development work that led to the first successful hybrid-type telephone repeaters, Gherardi and Jewett[3] briefly described the condition for stability of the series-type negative repeater. Later, in 1931, Crisson[4] considered both series- and shunt-type negative impedance repeaters and presented a detailed account of the conditions that had to be imposed for realization of a stable circuit. The first practical negative impedance repeater was due to Merrill,[2] who developed a negative impedance electron tube repeater, the stability and operation of which could be accurately predicted.

4.3 *The E6 Repeater*

In the E6 repeater impedance-conversion, impedance-matching and gain-adjusting functions are, in contrast to the E23 repeater, allocated to separate portions of the repeater. All impedance-matching and impedance-adjusting features are accomplished in the line building-out networks and gain adjustment is relegated to the series- and shunt-type elements of the converter. As a result, the theory of operation of the line building-out networks or of the converter may be considered separately once the impedance of the terminating cable as seen through the LBO network has been specified.

4.3.1 *Representation of Cable and LBO Network by a Lattice*

In the E6 repeater, certain portions of the LBO network are used — as the name "line building-out" implies — to build out fractional end-sections of cable to full-section value. The remaining portions of the LBO network consist of high- and low-frequency correcting networks used to match the impedance of the built-out cable to that of the converter.

The impedance and loss-frequency characteristic of such built-out and impedance-transformed cable can be represented by an equivalent lattice network as is shown in Fig. 5. The E6 repeater, arranged for terminal use, is shown in the upper portion of the figure. The image impedances appearing at each junction of the circuit are indicated. Considering only that portion of the circuit between the vertical broken lines, it can be seen that the LBO network, loaded cable facility and impedance compensator (nonrepeatered LBO network) have been replaced by an equivalent lattice having variable series and shunt impedance arms. Expressions defining the image impedance and insertion loss of an

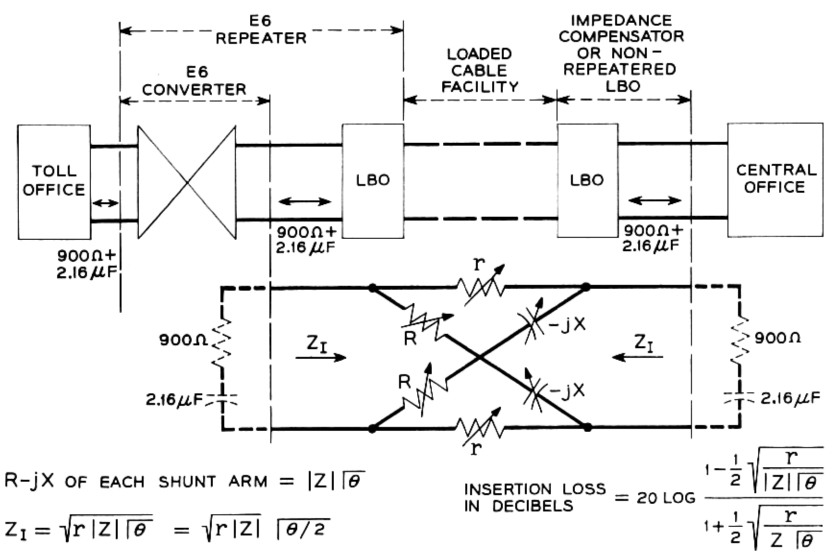


Fig. 5 — Equivalent lattice network of built-out section of loaded cable.

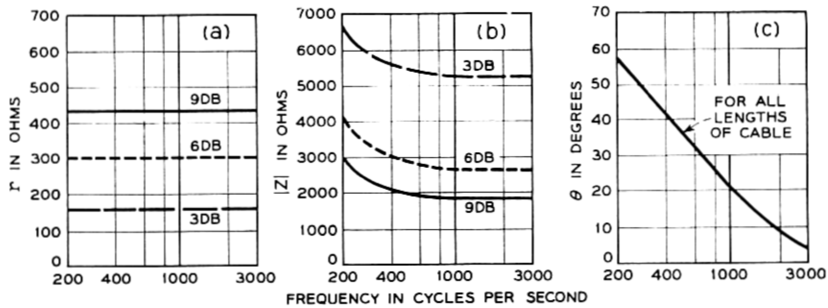


Fig. 6 — Approximate values of r and $Z \angle\theta$ for built-out 19-gage H-88 cable.

equivalent lattice are also shown. Although an impedance compensator, or nonrepeatered LBO network, is shown at the central office end of the trunk circuit, this network is usually omitted.

The magnitude of the series-arm impedance of the equivalent network of Fig. 5 increases with cable length, while the magnitude of the shunt-arm impedance decreases. Approximate values of the series- and shunt-arm impedances for lengths of cable having losses of 3, 6 and 9 db at 1000 cps are shown in Fig. 6. It is important to note that the angle θ specified for the shunt arm of the lattice is constant for all lengths of cable, varying only as a function of frequency.

4.3.2 *Lattice and Bridged-T Equivalents of E6 Converter*

Inasmuch as a lattice representation of the cable and LBO network has been defined, impedance requirements for both the series- and shunt-type converters of the repeater may now be established. In Fig. 5, assume that the E6 converter is replaced by an equivalent lattice. Assume further that the series and shunt arms of this lattice are equal in magnitude, but opposite in phase to those of the lattice equivalent of the built-out cable. Under these conditions, the image impedances presented by the converter will be maintained and the gain-frequency characteristic of the lattice will compensate for the loss-frequency characteristic of the built-out cable. The equivalent lattice network of the converter is shown in Fig. 7(a). Expressions for image impedance and insertion-gain are also shown.

Although the lattice representation of the E6 converter is a convenient tool for analysis, the bridged-T equivalent is more convenient for physical realization. In Fig. 7(b) the negative impedances required for the series and shunt arms of such a configuration are indicated. In Fig. 7(c)

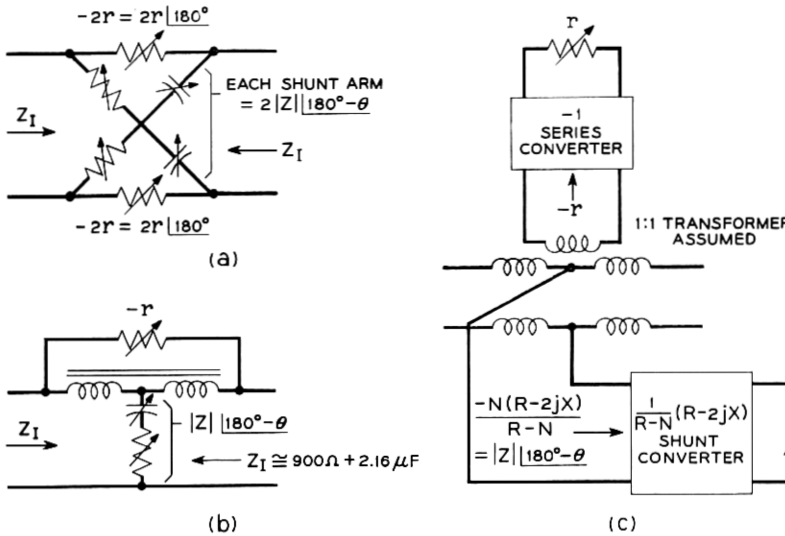


Fig. 7 — The E6 converter and equivalent circuits: (a) lattice network equivalent of E6 converter; (b) bridged-T equivalent of E6 converter; (c) E6 series and shunt converters.

the actual series and shunt converters are defined and shown in a balanced configuration employing a line transformer. It will be noted that the gain-adjusting networks associated with both converters are resistive. The series converter has a conversion factor of -1, while the shunt converter requires a conversion factor of $-\ N(R\ -\ 2jX)/(R\ -\ N)$. These conversion factors will be derived in discussion of the idealized E6 converter.

A more detailed schematic of the E6 converter is presented in Fig. 8. Only those circuit elements necessary for an understanding of converter operation are included. Furthermore, those elements which have no effect on idealized or theoretical operation are shown dotted. These include the simplified bias arrangements and transformers T2 and T3, which prevent longitudinal line voltages from disturbing the DC biases.

4.3.3 *Series Converter of the E6 Repeater*

The series-connected converter of the E6 repeater is, ideally, a four-terminal network having an input impedance that is the negative of the terminating resistances connected across its output. Such negative impedances, which are of the series or reversed-voltage type,[4] are open-circuit stable. Under certain restrictions that will be developed later, they may, therefore, be inserted in series with a transmission line to produce amplification without rendering the line self-oscillatory. Specifically, the term "open-circuit stable" indicates that the line terminals of the converter may be left open-circuited without causing the converter to sing.

The principle of operation of the E6 series converter is essentially the same as that of the impedance-terminated E2 electron tube repeater, or that of the transistorized E2 prototype described by Linvill.[5] The E6 converter, however, uses only resistance networks for gain control.

In order to describe the theory of operation of the series converter, the two-transistor arrangement previously shown in Fig. 8 will be employed. In the actual converter, as will be seen later, a double-compound transistor arrangement involving four transistors is used, but the simplification in no way invalidates the theory. Further simplification is

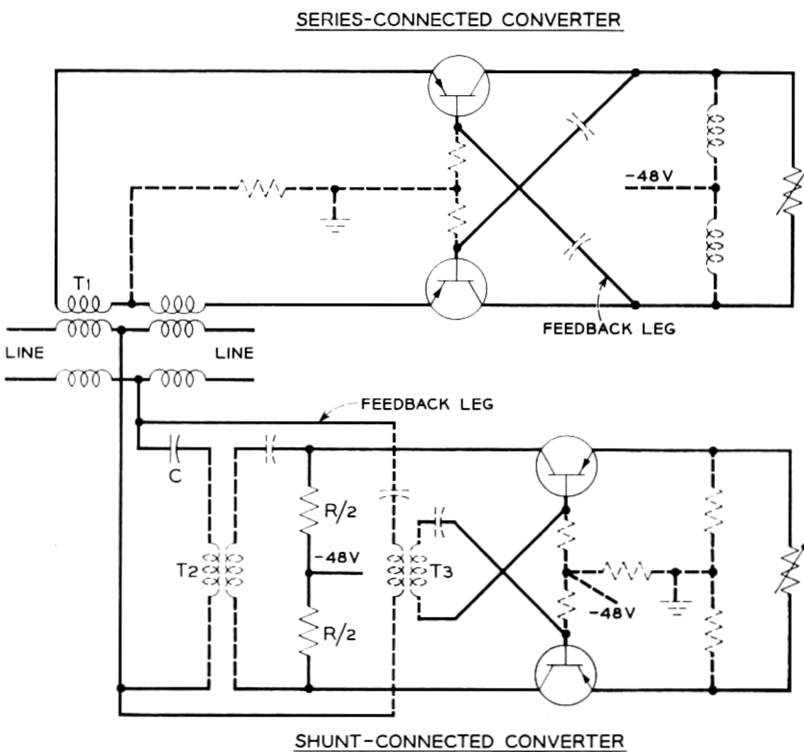


Fig. 8 — Simplified circuit of E6 converters with single transistors in each side of push-pull circuits.

achieved by neglecting biasing resistors, and by assuming that transistor alphas are equal to unity and that base and emitter resistances approach zero.

For the simplified schematics of Fig. 8, mesh equations could very easily be written and solved for the input impedance of the series or shunt connected converters. However, as soon as parasitic elements, cross-coupling or feedback capacitors and compensating or balancing elements are introduced, the mesh equations become unwieldy and difficult to interpret. If, however, a step-by-step description of the currents and voltages existing in the circuit is attempted, considerable simplification results.

In the usual mesh analysis, emitter and collector currents flowing in the circuit are assumed to have been established. Fig. 9(a) shows the direction of current flow in both emitter and collector circuits of a grounded-base transistor circuit. In Fig. 9(b), however, only the initial currents flowing in the emitter circuits of the series-connected converter

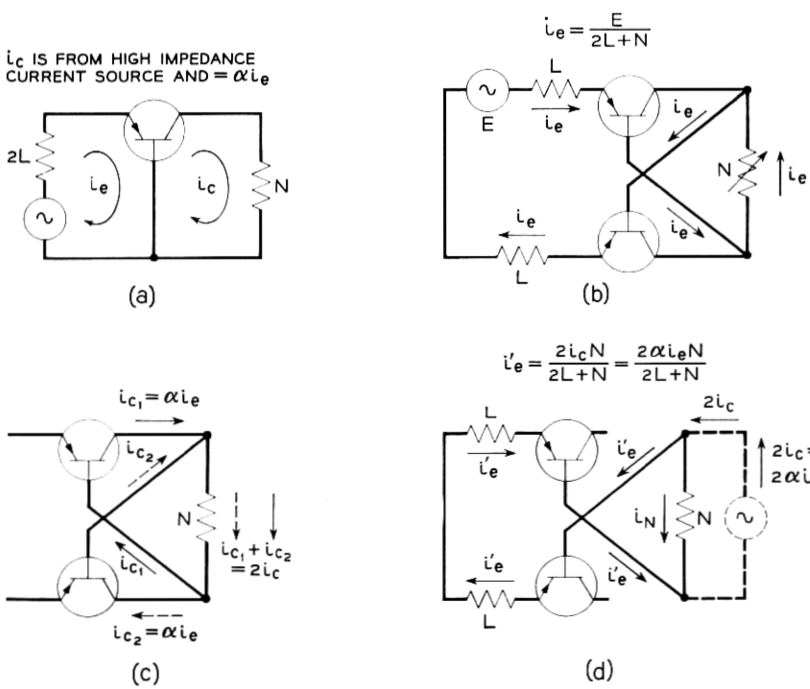


Fig. 9 — (a) Method of showing transistor currents; (b) emitter current due to line voltage E; (c) collector currents due to emitter current of (b); (d) positive feedback path completed through lines $2L$ to give i_e'.

due to application of the line voltage E are considered. As these currents flow, transistor action sets up collector currents in the collector portion of the circuit. In Fig. 9(c), only the collector currents flowing in circuit are shown. Each collector contributes a current equal to αi_e, and these combine in the adjustable resistive network N to give a total current of $2i_c$ or $2\alpha i_e$.

Considered from a slightly different point of view, the network of Fig. 9(c) may be replaced by that of Fig. 9(d), in which the current $2i_c$ is derived from an infinite impedance current source. In Fig. 9(d) the positive feedback path existing in the series-connected converter is completed through the line impedances, and the current $2i_c$ divides between these line impedances and the network N in inverse proportion to their magnitudes. Thus,

$$i_e' = \frac{2i_cN}{2L + N} = \frac{2\alpha i_eN}{2L + N}.$$

This current is in phase with the original emitter current. Since the original current was i_e, the $\mu\beta$ of the series converter may be defined as:

$$\mu\beta = \frac{i_e'}{i_e} = \frac{2\alpha N}{2L + N}.$$

For $\alpha = 1$,

$$\mu\beta = \frac{2N}{2L + N}.$$

With the completion of the feedback path as described above, the effective currents flowing in the converter may now be specified. These currents are shown in Fig. 10(a). The subscript T added to i_e and i_c indicates that these currents are due to the total effect of both the line voltage E and transistor and feedback action. Equal and opposite currents in both the network N and in the feedback legs cancel. The net current flow in the feedback or base-connected branches is reduced to zero, while the current in the network N is effectively in the reverse direction to the current originally established by the emitter alone.

If the net current in a feedback leg is zero, the end-points of the feedback branch must be at the same potential. Thus, in Fig. 10(b), A and B are at the same potential and, similarly, points C and D are at the same potential. Since this is the case, the net effect of feedback and transistor action in the converter is to establish the current and voltage relations shown in Fig. 10(b). If the voltage drop across network N is V, the voltage across the input terminals of the converter is also V, but is reversed

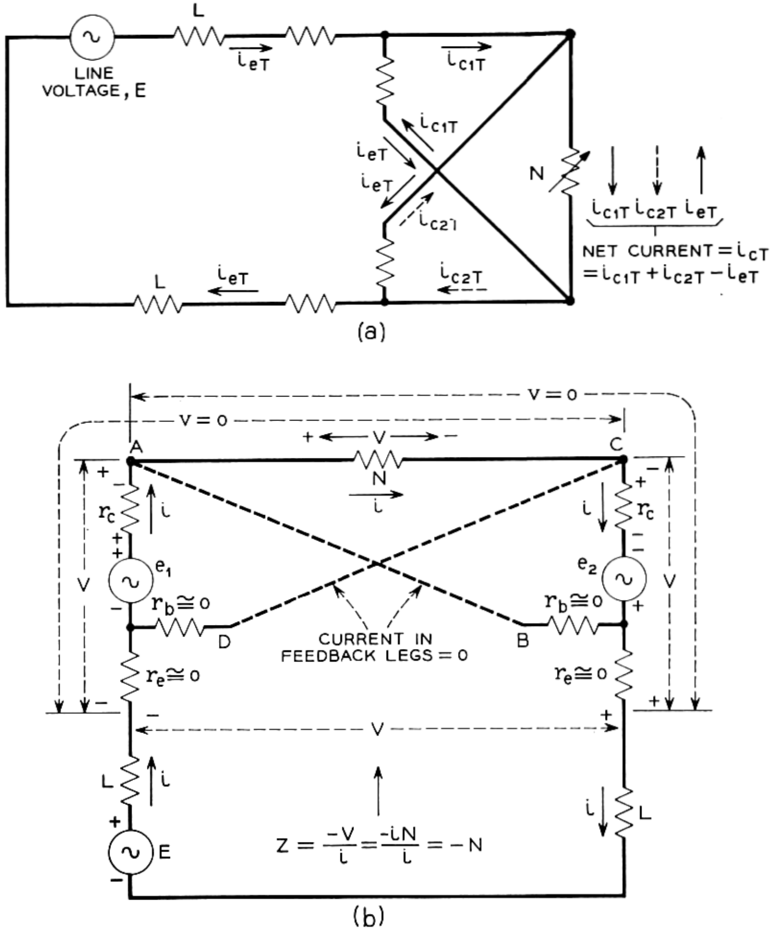


Fig. 10 — (a) Total currents in series converter due to line voltage and feedback; (b) voltages and currents in active series converter.

in sign. This voltage is in series with the line and, hence, the name "series or reversed voltage" type of converter.

In order for the voltages around the emitter-collector of the circuit to meet the requirements of Kirchoff's law, the four voltages indicated around the periphery must occur with the magnitudes and polarities as shown. The polarities of the voltages e_1 and e_2 in the collector circuits control the polarities of the voltage V across the transistors, since the effective collector voltage drop is small due to positive current feedback.

Considering the foregoing description of circuit operation from the standpoint of summing voltages around the circuit loops of Fig. 10(b), the input impedance of the series converter may be written by inspection, i.e.,

$$Z = -\frac{V}{i} = \frac{-iN}{i} = -N.$$

This impedance is closely approximated in the E6 series-connected converter for frequencies between 300 and 3000 cps and up to gains of approximately 13 db for the complete E6 repeater.

4.3.4 *Shunt Converter of the E6 Repeater*

Like the series converter, the shunt converter of the E6 repeater is a resistance-terminated four-terminal network. The shunt converter, however, does not present a resistive load to the line, but instead presents a complex impedance that is derived from a fixed phase-shift network included within the converter.

The shunt converter, as contrasted to its series counterpart, is of the shunt or reversed-current type.[4] It may, therefore, be placed in shunt across a transmission line without rendering the line self-oscillatory. Such a converter is short-circuit stable and, as a result, the input terminals of the device may be short-circuited without causing the converter to break into oscillation.

The basic principle of operation of the shunt converter is very similar to that of the series type. It should be noted, however, that the negative impedance of the shunt converter is obtained at the collectors of the converter rather than at the emitters. This is indicated in Fig. 11.

In developing the theory of operation of the shunt converter, the same step-by-step approach used in the series case will be found of value. In this instance, however, the derived voltages, indicated by the symbol V and shown around the outer periphery of the shunt converter circuit of Fig. 11, include the voltage drops across the phase-shift reactances $(-jX)$ as well as those across the collectors. It will also be noted that the voltage V across the input terminals of the converter is of the same polarity as that of the voltage developed across L2. This is the reverse situation from that of the series case. Furthermore, the total current i_3 flowing out of the converter terminals is seen to be in the reverse direction from that caused to flow by application of the voltage E to the line. This gives rise to the designation "reversed-current negative impedance".

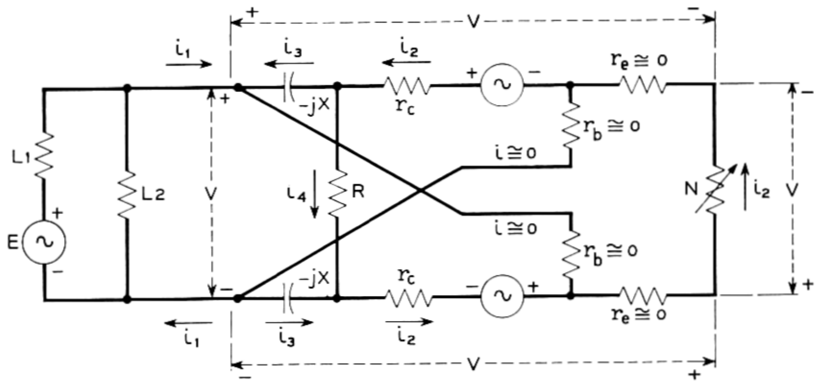


Fig. 11 — E6 phase-shift shunt converter.

As in the series case, the currents flowing in the feedback branches of the converter cancel and the relationship $V = i_2N$ can be written immediately. The other relationships needed for derivation of the negative impedance presented at the converter terminals follow from inspection of Fig. 11:

$$\begin{aligned} i_2 &= i_3 + i_4 , \\ V &= i_3 2jX + i_4 R, \\ i_3 &= -i_1 . \end{aligned}$$

Solving,

$$i_1 = \frac{V(R - N)}{-N(R - 2jX)}.$$

Since $Z = V/i_1$, the expression for the converter input impedance follows:

$$Z = \frac{-N(R - 2jX)}{R - N}.$$

This can also be written

$$Z = \frac{-N(R - 2jX)}{R - N} = |Z| \quad \underline{|180° - \theta}.$$

Were it not for the phase shifter, the expression for Z would be merely

$$Z = -N.$$

The more complex relationship, obtainable with the phase shifter, represents the desired impedance for the shunt converter of the E6 repeater. It is evident from inspection of this equation that the phase angle of Z is dependent solely on the magnitudes of R and $-2jX$ and is in no way dependent on N. The magnitude of N can thus be varied at will to adjust repeater gain without effect upon the phase angle of the resulting negative impedance. The impedance Z is closely approximated by the E6 shunt converter for gains as high as approximately 13 db for the complete E6 repeater.

The phase shifter of the shunt converter controls the impedance of the E6 repeater and permits impedance matching of the repeater to the impedance of the line as modified by the LBO network. The phase shifter also controls the gain-frequency characteristic of the repeater as a function of gain setting. These features of the phase-shift network become apparent when the series and shunt converters are combined in the standard bridged-T configuration.

If the input impedance of the series-connected converter is denoted as Z_A and that of the shunt converter as Z_B the following relations may, as has been previously indicated, be written:

$$Z_A = -N_1 \underline{|0^\circ} = N_1 \overline{|180^\circ},$$

$$Z_B = \frac{-N_2}{R - N_2}(R - 2jX) = \frac{-N_2}{R - N_2}(R - jX)$$

$$= \frac{N_2}{R - N_2}\sqrt{R^2 + X^2} \quad \underline{|180 - \theta_2},$$

where

$$\theta_2 = \tan^{-1}\frac{X}{R}.$$

Since the image impedance of bridged-T connected negative impedance converters is specified by the relation

$$Z_I = \sqrt{Z_A Z_B},$$

substitution of the derived expressions for Z_A and Z_B results in:

$$Z_I = \sqrt{\frac{N_1 N_2 \sqrt{R^2 + X^2}}{R - N_2}} \quad \underline{\left|\frac{-\theta_2}{2}\right.}.$$

The image impedance of the E6 converter must approximate 900 ohms in series with 2.16 microfarads. This latter impedance may be expressed as:

$$Z_0 = R_0 - jX_0 = \sqrt{R_0^2 + X_0^2} \quad \underline{\left| \frac{-\theta_0}{2} \right.},$$

where

$$R_0 = 900,$$

$$X_0 = \frac{10^6}{2.16\omega} \text{ ohms}$$

and

$$\theta_0 = \tan^{-1}\left(\frac{X_0}{R_0}\right).$$

For Z_I to equal Z_0, the following equalities must hold:

$$R_0 = \sqrt{\frac{N_1N_2R^2 + X^2}{R - N_2}}$$

and

$$\tan^{-1}\left(\frac{X_0}{R_0}\right) = \tan^{-1}\left(\frac{X}{R}\right).$$

Except for the X^2 term appearing under the radical in the equality involving R_0, these relationships would be exact, and X^2 is small compared to R^2 at frequencies above 300 cycles. The degree of approximation that may be obtained is dependent on the proportion of N_1, N_2, R and X. These same factors determine the gain and gain-frequency characteristics of the converter and, as a result, both impedance and gain requirements must be considered in the selection of design values.

In order to derive an expression for insertion gain of the E6 converter, reference is again made to the general bridged-T network for which

$$\text{insertion gain (in db)} = 20 \log_{10} \left| \frac{1 - \sqrt{\dfrac{Z_A}{4Z_B}}}{1 + \sqrt{\dfrac{Z_A}{4Z_B}}} \right|.$$

Considering only the terms under the radical sign, and substituting expressions for Z_A and Z_B,

$$\sqrt{\frac{Z_A}{4Z_B}} = \frac{1}{2}\sqrt{\frac{N_1(R - N_2)}{N_2\sqrt{R^2 + X^2}}} \; \underline{|\theta_2/2} = M \; \underline{|\theta_2/2}$$

Insertion gain may, therefore, be written

$$\text{insertion gain (in db)} = 20 \log_{10} \left| \frac{1 - M \underline{|\theta_2/2}}{1 + M \underline{|\theta_2/2}} \right|$$

Taking the real and imaginary parts of $M \underline{|\theta_2/2}$, this expression becomes:

$$\text{insertion gain (in db)} = 20 \log_{10} \left| \frac{\left[1 - M \cos\left(\frac{\theta_2}{2}\right)\right] - jM \sin\left(\frac{\theta_2}{2}\right)}{\left[1 + M \cos\left(\frac{\theta_2}{2}\right)\right] + jM \sin\left(\frac{\theta_2}{2}\right)} \right|.$$

When the absolute value of this equation is taken, the expression for insertion gain reduces simply to:

$$\text{insertion gain (in db)} = 20 \log_{10} \sqrt{\frac{1 - \delta}{1 + \delta}},$$

where

$$\delta = M\left[M + 2 \cos\left(\frac{\theta_2}{2}\right)\right],$$

$$M = \frac{1}{2}\sqrt{\frac{N_1(R - N_2)}{N_2\sqrt{R^2 + X^2}}}$$

and

$$\theta_2 = \tan^{-1}\left(\frac{X}{R}\right).$$

As indicated in this equation, the over-all gain of the bridged-T connected converter is dependent on the gain settings of the series- and shunt-connected units (N_1 and N_2) and on the phase-shift parameters R, X and θ_2.

The capacitors in the phase shifter, plus others inserted in the feedback branches of the converter, also serve to increase the impedance of the converter to low frequency signals and dial pulses. By this technique, the inclusion of the shunt converter across the lines does not seriously decrease signaling ranges of most types of signaling systems used in Bell System exchange circuits.

4.3.5 *Comparison of Physical Converter Circuit to Ideal*

In the preceding discussions on principles of operation, the series and shunt units were first treated separately and then combined in a bridged-

T arrangement. The biasing and balancing networks arrangements were, however, omitted for simplification of presentation. These omissions are included in the complete schematic of the E6 converter shown in Fig. 12. The double-compound transistor arrangements used in both the series and shunt converter units are also indicated.

The principal assumptions made in the theoretical circuit descriptions were that transistor alphas were equal to unity and that the direct current biasing circuits could be neglected. In the following discussion, the

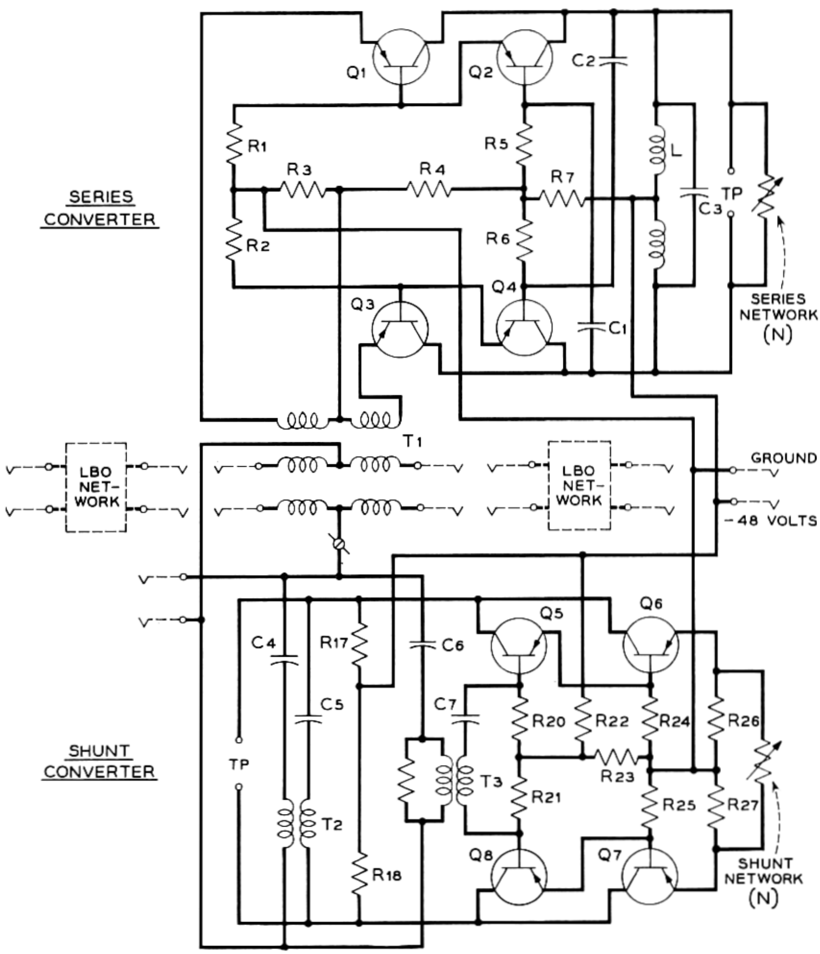


Fig. 12 — Complete schematic of E6 repeater.

validity of these assumptions will be assessed and the effects of other circuit elements on converter operation will be evaluated.

The double-compound transistor arrangement due to Darlington[6] has the advantage that the alpha of the compound transistor approaches unity more closely than the alpha of a single transistor. In addition, the use of such an arrangement insures that the over-all gain of the circuit is kept substantially constant regardless of normal variations in the individual transistor amplification factors. Each compound transistor, nevertheless, functions substantially as a single transistor and can be treated as such.

A more detailed sketch of the compound arrangement, corresponding to a single transistor, is shown in Fig. 13. Here the currents in the emitter, base and collector of each component transistor are defined. The compound alpha may be expressed as

$$\alpha_c = 1 - (1 - \alpha_1)(1 - \alpha_2).$$

Thus, if both transistors have a minimum alpha of 0.94, the compound alpha is found to be $\alpha_c = 0.9964$.

The DC bias arrangements of both the series and shunt elements of the converter are such as to have little effect on the angle of the conversion factor. Any changes in the real part of the conversion factor are small and can be compensated for in the series and shunt gain-controlling networks.

In the design of the E6 converters, the transformer T1 of Fig. 12 is considered as part of the series element of the converter. The purpose of this transformer is to combine the series and shunt elements of the converter in a bridged-T arrangement. The design of the transformer is quite critical in that the magnitudes of the leakage and mutual inductances and of the distributed capacitance parameters limit the maximum gain available from the series converter. The transformer design used

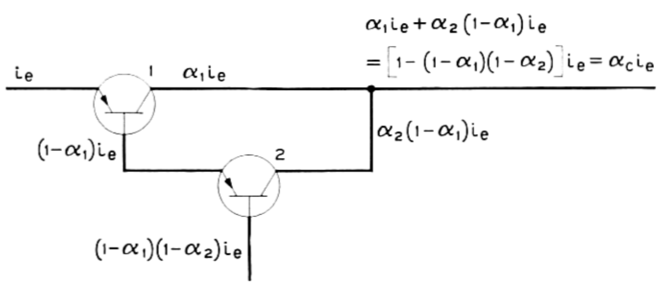


Fig. 13 — Compound transistor arrangement corresponding to single transistor.

in the E6 repeater, however, is not limiting up to gains of 13 or 14 db. It must, however, be capable of carrying the line currents due to talking battery or to signaling.

In the shunt-connected converter transformers T2 in the phase-shift network and T3 in the feedback circuit are used to isolate the DC bias arrangement from the effects of power-induced longitudinal voltages. In a small percentage of cables in the telephone plant, these 60-cps longitudinal voltages may be as high as 50 volts. The high mutual inductances and low series resistances of these transformers cause the transformers to have negligible effect in the transmitted band of the repeater. The use of these transformers also eliminates the longitudinal balance problem in the shunt connected converter. If the transformers were not specified, it would be necessary to introduce highly balanced capacitors in the tip and ring portions of the phase shifter and feedback circuits.

4.4 *Line Building-Out (LBO) Networks*

4.4.1 *General*

In a previous section it was pointed out that the impedance-matching properties of LBO networks were defined once the input impedance of the associated converter had been specified. Since the input impedance of the E6 converter is nominally 900 ohms in series with 2.16 microfarads (900 + 2.16 microfarads), LBO networks are required to match the characteristic impedance of variable end-section loaded cable facilities (of particular gage and type of loading) to this impedance.

A somewhat similar problem, which has been of interest to network designers for a number of years, is that of the broadband matching of an arbitrary impedance to a pure resistance. Although this general problem will not be considered in detail, the problem of matching the characteristic impedance of loaded cables to a pure resistance will be treated at some length. This is done in part for simplicity of presentation and to outline procedures that may be of practical value in the general case.

4.4.2 *Characteristic Impedance of Loaded Cables*

The characteristic impedance of loaded cable is not arbitrary, but is well-defined. It is, nevertheless, quite complex, varying with frequency, length of end-section, loading and cable conductor size. The characteristic impedance for one-half end-section 19- and 22-gage H-88 loaded cable is shown in Fig. 14. The mutual capacitance of these cable pairs is 0.082 microfarad per mile, and the frequency of cutoff is about 3480 cps.

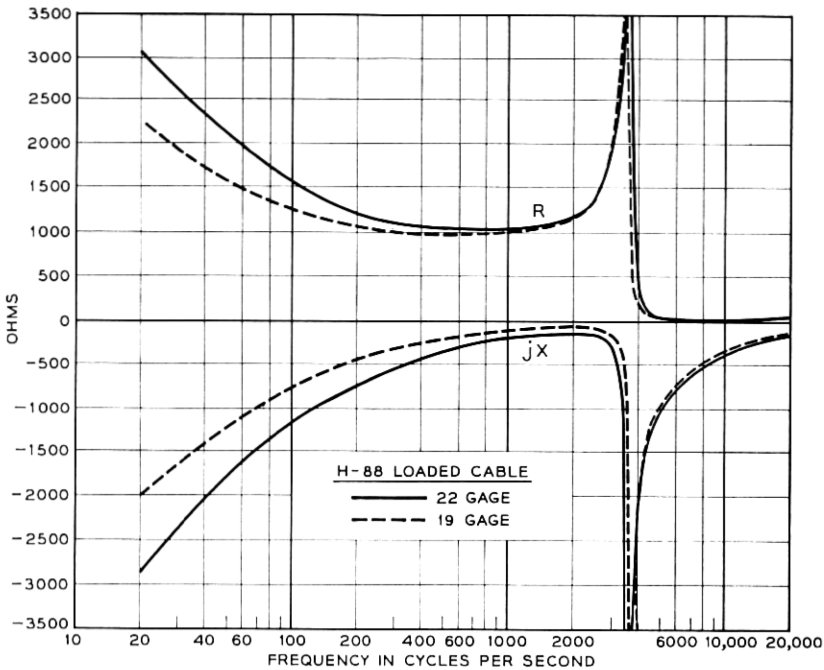


Fig. 14 — Characteristic impedance of 0.5 end-section H-88 loaded cable.

Not only does the impedance of these cable facilities vary with conductor size, but it varies with the length of end-section as well. The variation in impedance of 24-gage H-88 with varying length of end-section is shown in Fig. 15. The impedance characteristic below cutoff is indicated as the length of end section is increased from zero (adjacent to a loading coil) to full section (6000 feet from a loading coil).

The large variations in impedance shown in Figs. 14 and 15 complicate the job of matching the image impedance of the repeater to the characteristic impedance of the loaded cable. When it is specified that the matching be effective through and beyond cutoff, the situation is made even more difficult.

4.4.3 *Broadband Impedance Matching*

A broadband impedance-matching network is generally a two-terminal pair network similar to that of Fig. 16 and having the property that the input impedance Z_{11} approximates a pure resistance R when the network is terminated in an arbitrary impedance Z_x. For the realization of such a network, several design methods are available; however, an

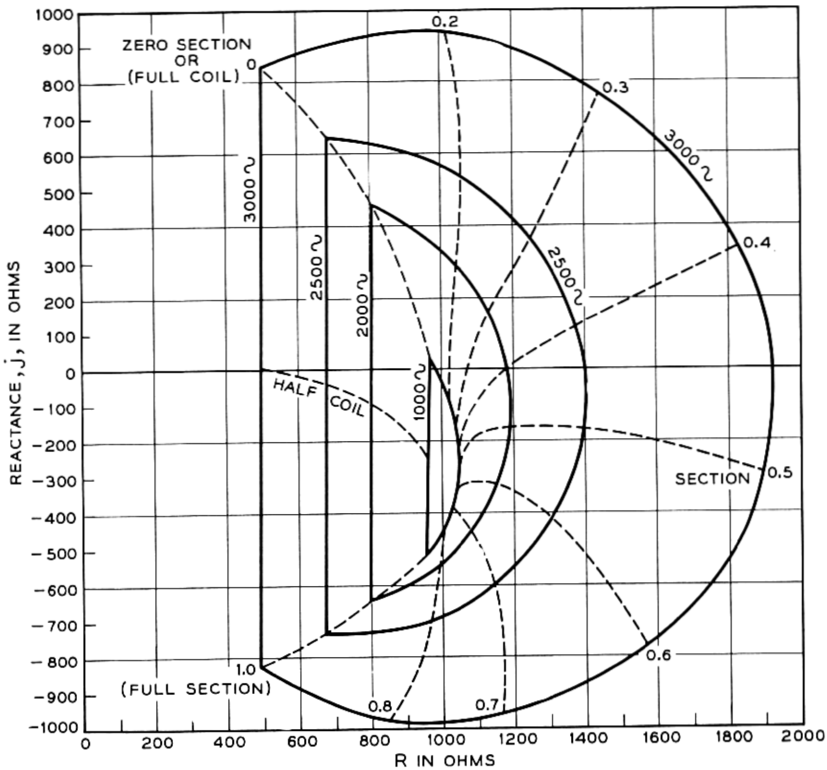


Fig. 15 — Impedances at various frequencies and variable end-sections for 24-gage H-88 loaded cable.

optimum network for a particular application depends on the impedance to be matched, the maximum tolerable flat loss and other requirements.

The case of lossless matching presents a continuing challenge. Basic limitations in terms of bandwidth and reflection coefficient are given by Fano,[7] but the technique is limited to a few special types of arbitrary impedance and is tedious to calculate. The design technique presented by Fano has been modified by Matthaei[8] to give straightforward proce-

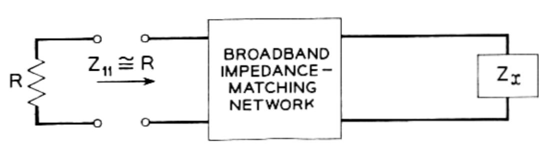


Fig. 16 — General network arrangement for matching an arbitrary impedance to a pure resistance.

dures for synthesis of optimum, lossless, Tchebycheff, wideband impedance-matching networks for various classes of loads. Another approach requires an approximation of the conjugate of the given load impedance and its realization as a Darlington-type network.[9] In general, this latter approach also results in networks that are both complicated and expensive to build.

The techniques of lossy matching, however, are relatively simple. In Fig. 17 three methods of design are indicated. If Z_x is the given arbitrary impedance and Z_a is a rational approximation to Z_x, then it is evident that all three networks satisfy the matching requirement. These networks, although deceptively simple, may in practice become quite complicated. Furthermore, the networks of Fig. 17(a) and (b) will be excluded from consideration in many practical applications due to the inherent 6-db flat loss of the networks.

The network of Fig. 17(c) would also appear to offer considerable difficulties in practice. Obviously, R has to be larger than the maximum real part of Z_a and, in addition, the realization of the $(R - Z_a)$ function could reasonably be expected to be both tedious and complex. In the general case, this is undoubtedly true, but, as will be shown later, this network results in a simple and effective solution to the broadband impedance matching problem when Z_x is taken to represent the characteristic impedance of loaded cable facilities.[10]

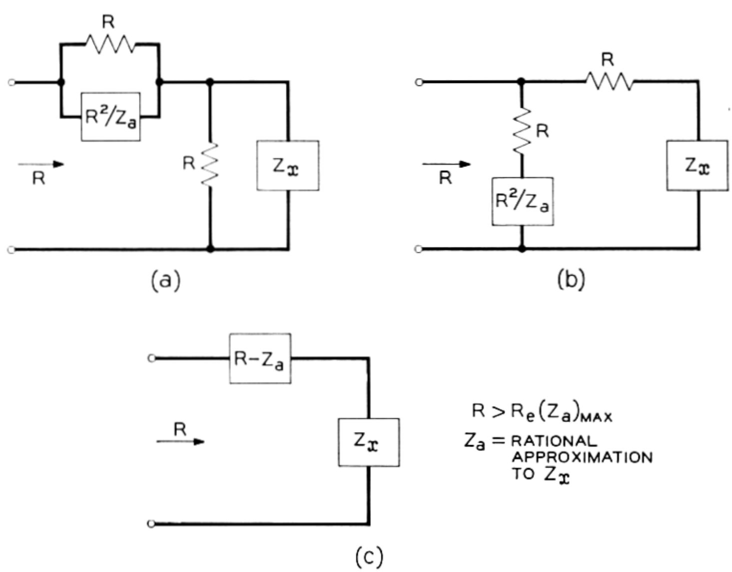


Fig. 17 — Lossy networks matching an arbitrary impedance to a pure resistance.

4.4.4 *Practical LBO Networks*

In the development of the E6 repeater, requirements placed on the LBO network precluded the use of any of the previously described methods of approach with the single exception of the method indicated in Fig. 17(c). Requirements on return loss, echo return loss and insertion loss were severe, but the added requirements on allowable size and cost were limiting, in that complex networks were thereby excluded. Requirements on low-frequency signaling margins were particularly severe, and these effectively established criteria for network configurations that might be employed. Other requirements relating to longitudinal unbalance, protection against high voltage surges, etc., also had to be considered.

The application of immitance-matching techniques to the design of LBO networks, however, has resulted in a number of practical networks. Three different approaches considered during the development are significant. These are: (a) the fractional termination method of image parameter theory; (b) the "lossy" matching method implied in Fig. 17(c) and (c) the method of impedance compensation.

4.4.4.1 *Fractional Termination Method.* This method of design is based on recognition of two essential facts: (a) a midsection loaded cable can be simulated approximately by a midshunt constant-K image impedance and (b) properly terminated constant-K low-pass and high-pass filters having identical cutoff frequencies, operated in parallel, are complementary.

Since the fractional termination method of conventional image parameter theory has been completely described in the literature,[11] only a brief outline of the theory will be attempted here.

In Fig. 18, a parallel combination of a constant K low-pass and high-

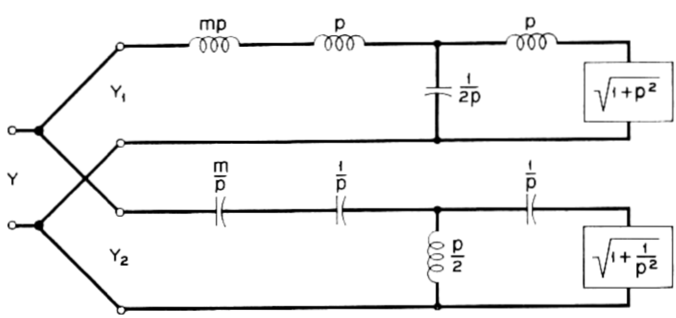


Fig. 18 — High-pass–low-pass fractional termination network.

pass filter is shown. The impedances and frequencies are normalized to unity. Thus,

$$Y_1 = \frac{1}{mp + 1 + p^2},$$

$$Y_2 = \frac{p}{m + 1 + p^2}.$$

In the low-pass band, $p = j\omega$, $\omega \leqq 1$; hence,

$$Y_1 = \frac{1 - \omega^2}{m^2\omega^2 + 1 - \omega^2} - \frac{jm\omega}{m^2\omega^2 + 1 - \omega^2},$$

$$Y_2 = \frac{j\omega}{m + 1 - \omega^2}.$$

Similar expressions for the high-pass band can be found. The real and imaginary parts of Y_1 and Y_2 are plotted in Fig. 19. It is recognized that

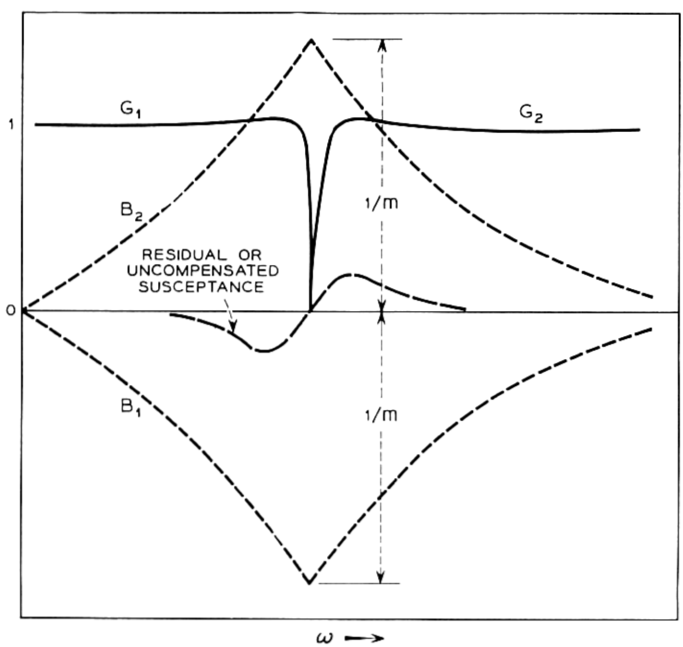


Fig. 19 — Conductance and susceptance characteristics of fractional termination network.

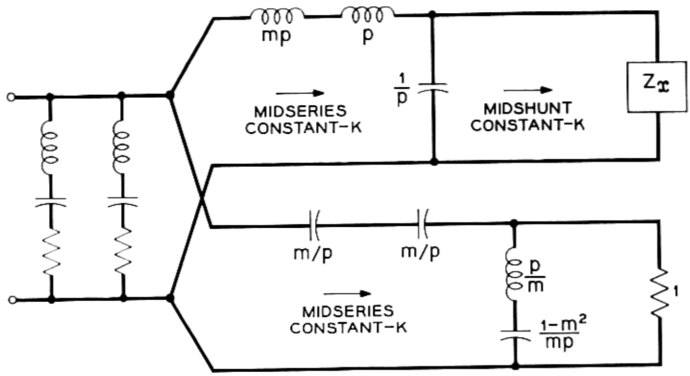


Fig. 20 — Fractional termination network with image impedances replaced by physically realizable terminations.

the real parts are the m-derived image impedances; hence, the input admittance Y is, roughly, a constant resistance except for residual effects in the vicinity of cutoff. The matching properties of the network may be improved by the addition of a simple series RLC susceptance-correcting network across the input terminals.

To apply the theory just described, the network shown in Fig. 20 is designed. A constant-K half section is first built out from the cable to form a mid-series impedance so that a parallel combination of low- and high-pass filters can be formed. A susceptance-correcting network, as described above, is added across the input terminals to correct for residual effects resulting from the formation of the complementary filters. A second series RLC correcting network is also added to correct for the deviations of the cable impedance from that of a constant-K low-pass image impedance at low frequencies. Fig. 21 shows the final configuration of the fractional termination type of matching network.

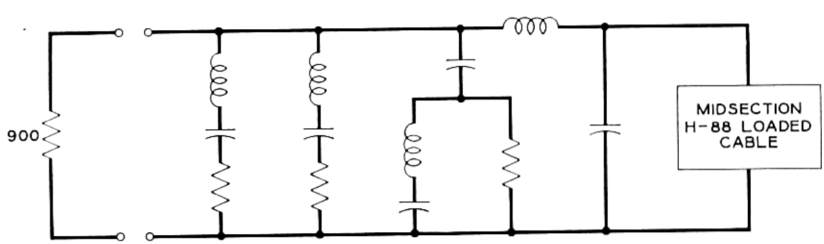


Fig. 21 — Two-terminal pair fractional termination LBO network.

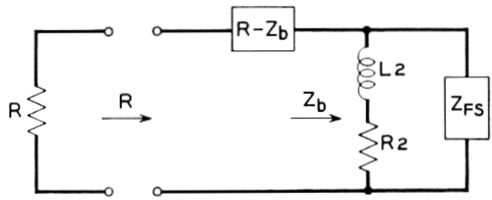


Fig. 22 — Lossy matching of full section of loaded cable to a pure resistance.

4.4.4.2 *Lossy Matching.* Although the fractional termination method of design results in practical networks, these networks are too complex from a cost point of view. In addition, the networks are limited to the extent that the cable impedance can be built out to represent a midshunt constant-K image impedance.

One of the simplest approaches to the design of LBO networks is based on the realization of a network that, when placed in series with the cable impedance, results in a pure resistance match. Such a network, as shown in Fig. 17(c), was described in the general discussion of lossy matching.

In the network of Fig. 17(c), Z_a is a rational approximation to the terminating impedance Z_x. Except for the criterion that R be larger than the maximum real of Z_a, no restrictions are placed on the approximating function $(R - Z_a)$. As previously indicated, the realization of such an approximating impedance function might be quite difficult. In the case of LBO networks for loaded cable facilities, however, the synthesis problem is quite simple, provided certain impedance correcting techniques applicable to loaded cable are recognized.

In Fig. 22 the generalized impedance Z_x is replaced by Z_{FS}, which represents the characteristic impedance of a full section of loaded cable. Except for the specification of a full section of cable, this substitution is quite obvious. It will be noted, however, that Z_{FS} has been shunted by a simple series RL network, with the impedance of the parallel combination being indicated as Z_b. The technique of lossy matching obviously calls for the addition of a series network having the impedance $(R - Z_b)$. Unless the characteristic impedance of loaded cable were investigated, the above procedure would indicate a further complexity in design. However, as shown in Fig. 23, a considerable simplification results, for the function $(R - Z_b)$ can now be approximated by a simple series RC network in parallel with an inductor.

Assuming that the simple network described above is capable of yield-

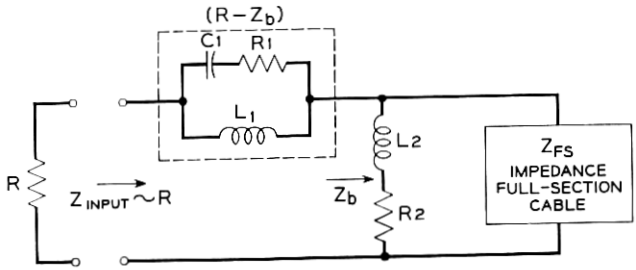


Fig. 23 — Lossy matching approximation to synthesis of LBO networks.

ing satisfactory results, it would appear that certain difficulties still exist. The network requires a full-section cable termination. Fortunately, the impedance of variable end-sections of cable can be easily built out to full-section value by the addition of series resistances and shunt capacitances.

Although the network of Fig. 23 is capable of meeting most of the requirements specified, it cannot be used without modification because of the signaling penalities that result from its application. The series RL network connected across the full-section cable impedance shunts direct currents, presents a low impedance to signal frequencies and degrades service. By placing a sufficiently small capacitor in series with the series RL network, the signaling penalties can be reduced to any desired extent, but the effectiveness of the admittance correcting network R_2L_2 is correspondingly impaired.

A compromise must, therefore, be reached. For signaling purposes it is possible to specify a minimum value of capacitance that can be tolerated. Once this has been done, a new network may be designed following the procedures previously outlined. Since the resulting network represents a compromise, degradation in certain characteristics must be expected. The complete network including the building-out capacitors, building-out resistors and the "signaling" capacitor is shown in Fig. 24.

4.4.4.3 *Impedance Compensation Method*. In order to describe the impedance compensation method adequately, it is necessary to reconsider the variations that take place in cable impedance as the length of end-section is varied. In this instance, it is more convenient to plot the real and imaginary components of the cable impedance as functions of frequency.

The wide variations that take place in the characteristic impedance of loaded cable as the length of end-section is varied is indicated by the solid curves of Figs. 25, 26 and 27 for 0.2, 0.5 and 1.0 end-section. It

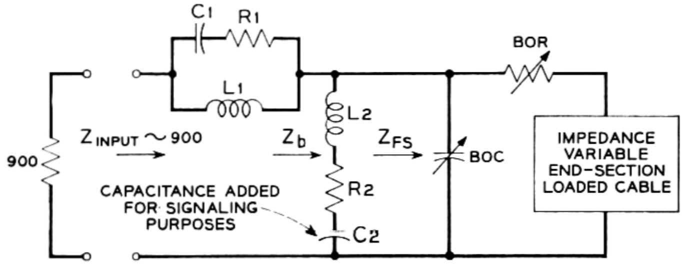


Fig. 24 — Practical LBO network for matching variable-end-section loaded cable to 900 ohms.

will be observed that the resonance effects in the vicinity of cutoff either disappear or become highly damped as the length of end-section is increased.

As noted previously, variable end-sections of cable may be built out to full-section value by the addition of series resistance and shunt capacitance. The effect of adding only capacitance to the fractional end-section is indicated by the same figures. It may be seen from comparisons between the fractional and full end-section characteristics that a series resistance must be supplied to raise the resistive component of a cable

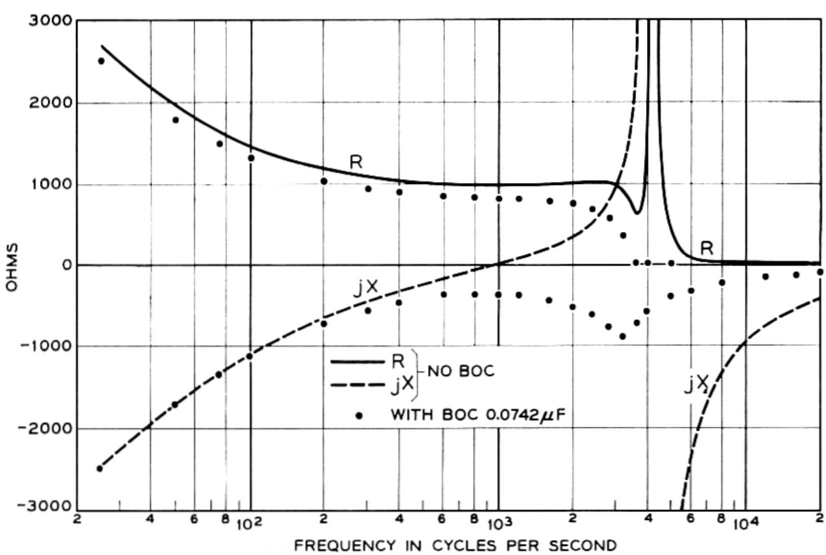


Fig. 25 — Characteristic impedance of 0.2 end-section 22-gage H-88 loaded cable with and without building-out capacitance network added.

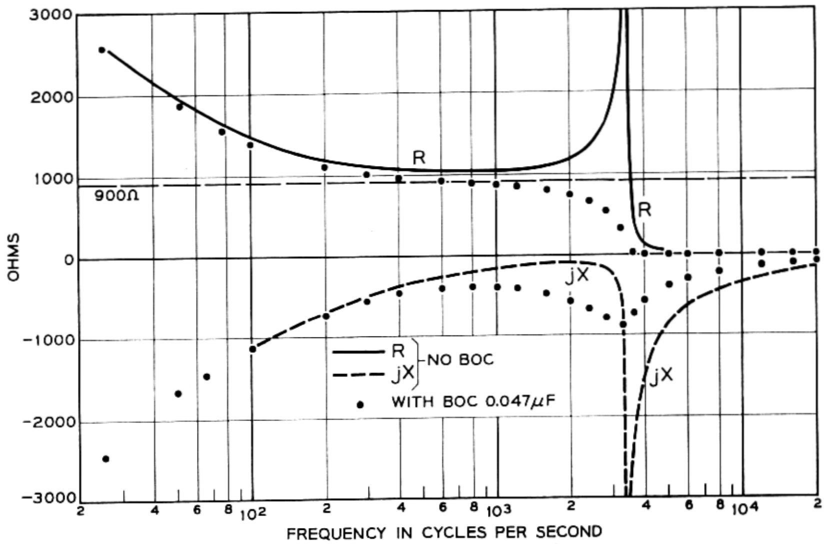


Fig. 26 — Characteristic impedance of 0.5 end-section 22-gage H-88 loaded cable with and without building-out capacitance network added.

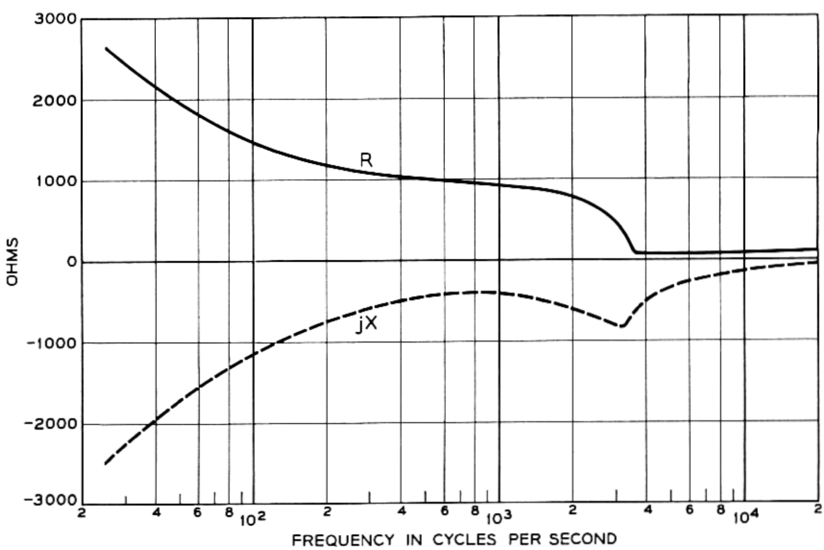


Fig. 27 — Characteristic impedance of full-section H-88 loaded cable.

built out with capacitance to the desired full-section value. In the final development model of the LBO network, the resistive elements supplied for building out purposes are designated as BOR (building-out resistance) networks. The capacitive elements are designated as BOC (building-out capacitance) networks.

From a study of the impedance characteristic of built-out cable, it may be seen that impedance-correcting techniques have application to the design of LBO networks. This becomes quite apparent if, as in Fig. 28, the impedance characteristic is divided into two zones of interest along the frequency axis. The first of these zones, the high-frequency correcting (HFC) zone, requires a series impedance type of correction. The second, the low-frequency correction (LFC) zone, requires a shunt impedance or admittance type of correction.

The problem of impedance correction in the high-frequency zone is easily solved by application of the simple network of Fig. 29(a). The schematic shows both the network and the general shape of its two-terminal impedance characteristic. In Fig. 29(b) the corrected impedance of the built-out cable is indicated for the idealized case. If the admittance characteristic of the high-frequency corrected cable impedance is now plotted, a characteristic similar to that of Fig. 30 results.

It is quite obvious that the design of the low-frequency correcting network will depend on the type of repeater with which the LBO network is to be associated. If a pure-resistance matching network is required, the low-frequency correcting network must completely annul the residual susceptance of Fig. 30. On the other hand, if a 900 + 2.16

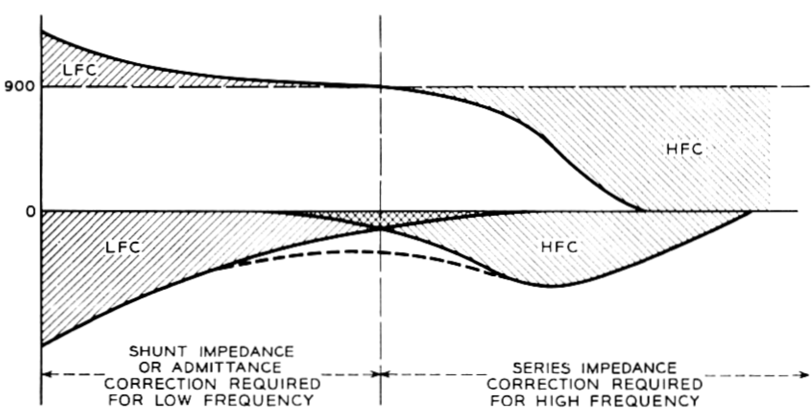


Fig. 28 — Division of LBO network design into low- and high-frequency correction zones.

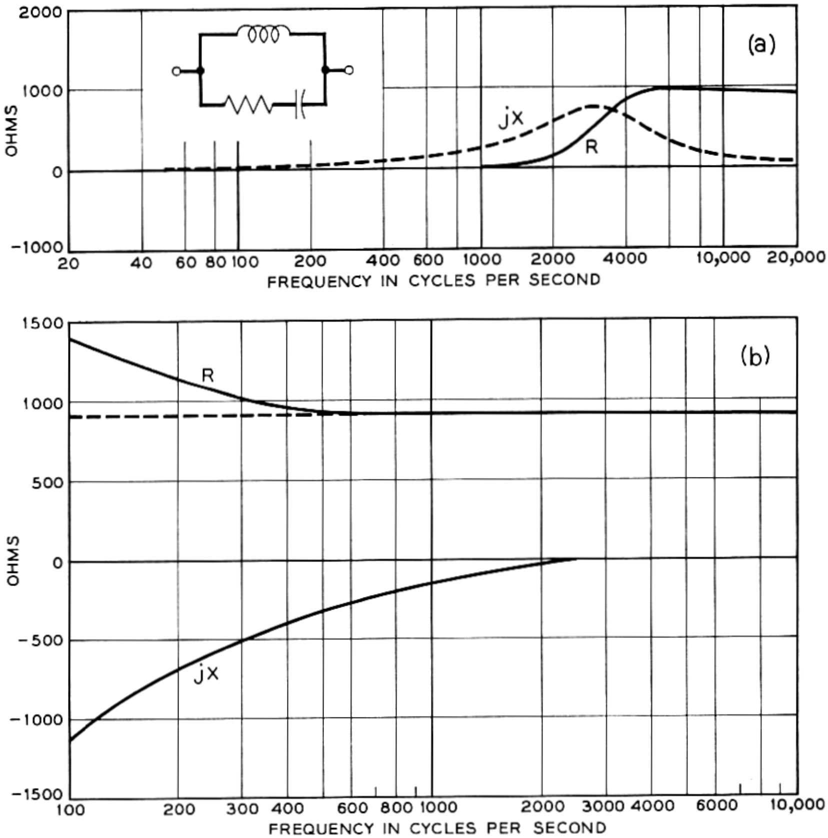


Fig. 29 — High-frequency impedance correction.

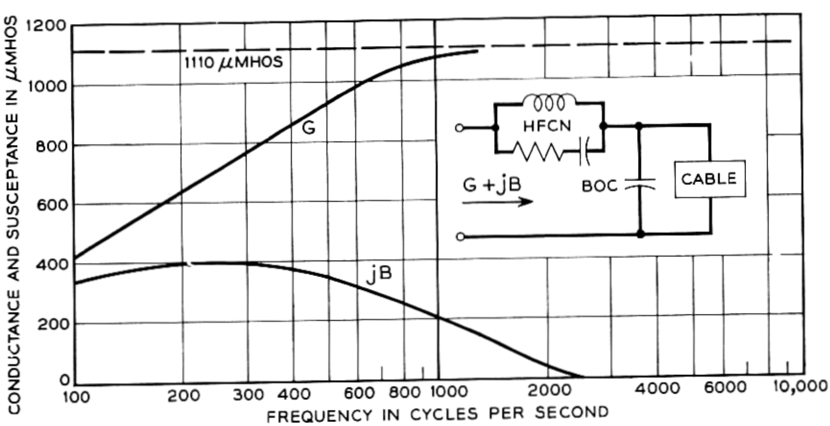


Fig. 30 — Admittance characteristic of high-frequency network, BOC, cable.

microfarad network is required, the susceptance need be annulled only partially. As previously noted, a series capacitor must be included for signaling and supervisory reasons.

4.4.5 *Description of Final Development of LBO Network*

In previous sections, the LBO networks were shown as unbalanced structures. Inasmuch as these networks must work with balanced cable pairs, however, the actual network requires balanced construction for the BOR and HFC networks.

As indicated in Fig. 31, the final development model of the LBO network (designated as the 830-type network) includes a resistance (BOR) and a capacitance (BOC) network for building out cable pairs to full-end section value. These networks are required for all gages of cable. The high-frequency correcting network represents a compromise in design suitable for all gages of cable. The three-winding transformer indicated in the sketch inserts the required balanced inductance in each series branch of the network and effectively places the single *RC* network termination of the third winding in shunt with each inductance. Problems of longitudinal balance are, therefore, minimized and only one *RC* network is required.

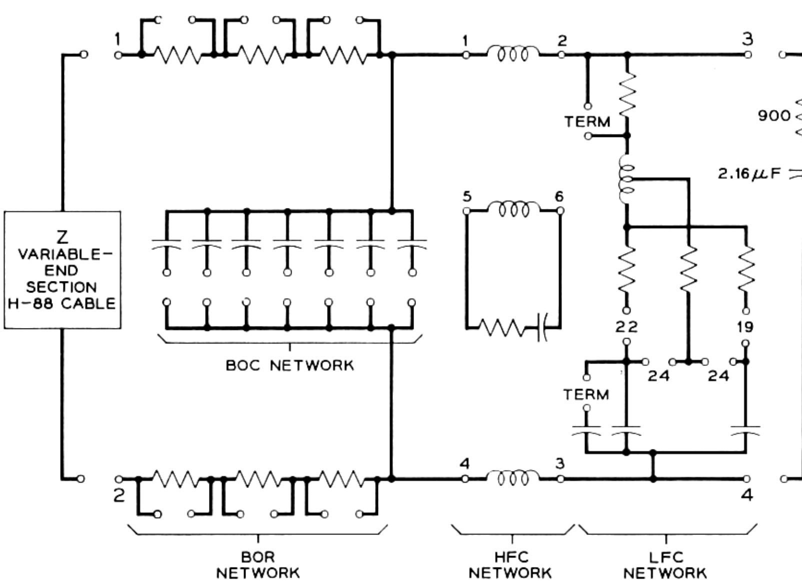


Fig. 31 — Complete line building-out network for H-88 cable facilities.

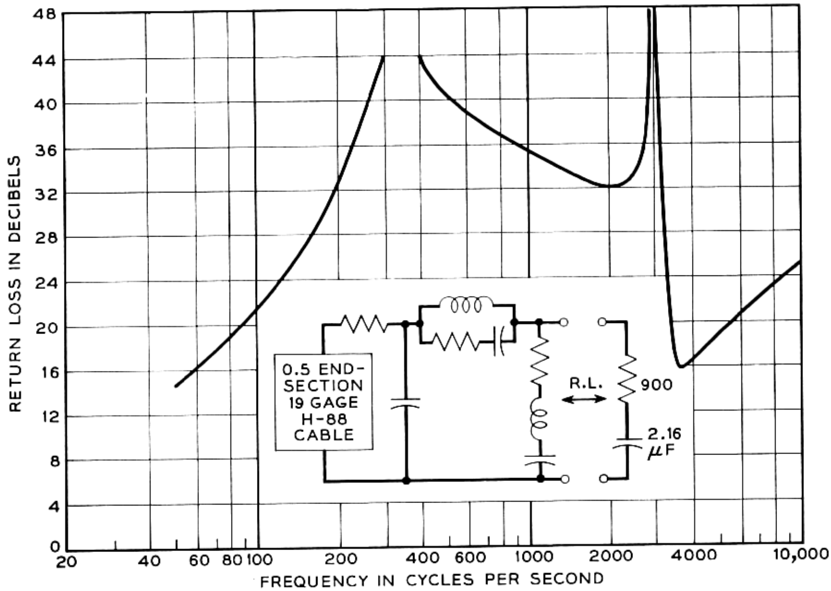


Fig. 32 — Return-loss characteristic of line-building-out network for 0.5 end-section 19-gage H-88 cable.

Although compromises were also required in the low-frequency correcting network section of the LBO network, separate LFC networks had to be specified for each gage of cable. The schematic of Fig. 31 also indicates how network elements are combined to minimize the total number of elements required.

The impedance-matching possibilities of this simple LBO network are indicated by the return-loss characteristic of Fig. 32. In the sketch, the element values are shown unbalanced. For matching 0.5 end-section 19-gage H-88 cable, no resistance is included in the BOR network.

Two LBO network designs were initially made available for systems use. The cable facilities that may be accommodated by the two LBO networks are shown in the application chart of Table I.

Table I — Cable Facilities Application Chart for LBO Network

LBO Network	Type of Cable	Gage	End-Section
830A	H-88 loaded, high capacitance	19,22,24	0—full
830B	H-88 loaded, low capacitance	19	0—full
830B	D-88 loaded, high capacitance	22	0—full

V. PERFORMANCE CHARACTERISTICS OF THE E6 REPEATER

5.1 *Impedance and Return Loss*

A typical example of the impedances existing at the converter-LBO network junction is shown in Fig. 33. The dashed lines show the impedance of 24-gage H-88 cable as modified by the LBO, and the solid lines show the impedance of a gain unit having a gain of 6 db. Curves for other cable facilities and other gains are very similar, because the combined converter impedance is quite constant with gain while the LBO networks correct for the differences between the several cable facilities.

A method for checking that the series and shunt converters have the correct relative impedance for any given gain setting is to make the individual 1000-cycle insertion gains of these converter units the same when they are each measured between image impedances of the converter. If this adjustment procedure is correct, the image impedance of the converter should result when the expression for the loss of the series-connected converter is equated to that of the shunt-connected converter.

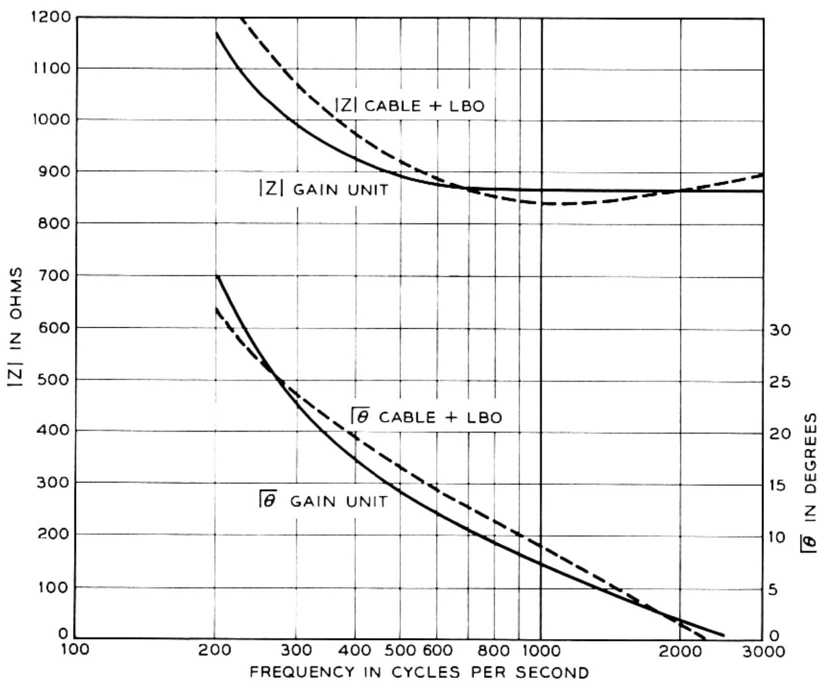


Fig. 33 — E6 gain unit impedance compared to impedance of 24-gage H-88 cable plus LBO.

In Fig. 34, a circuit arrangement used for individual series and shunt unit impedance adjustment by measurement of insertion gain is shown. Equating the two expressions indicated in Fig. 34,

$$20 \log_{10} \frac{2Z_I}{2Z_I - Z_A} = 20 \log_{10} \frac{-Z_B}{-Z_B + \dfrac{Z_I}{2}}.$$

Solving for Z_I, the familiar expression for the image impedance of the bridged-T negative impedance converter results:

$$Z_I = \sqrt{Z_A Z_B}.$$

The return loss between the E6 converter and the LBO and cable is about 34 db between 500 and 2500 cycles when the individual converter insertion gains are alike and the cable has nominal impedance. When the individual converter insertion gains differ by 0.2 db, this return loss is about 32 db, and when they differ by 0.4 db, it is about 28 db.

Tolerances on converter components are such that, when the over-all gain of the repeater is 6 db and the individual converter gains are set in accordance with tabulated instructions, these gains will not differ by more than about 0.2 db in 99 units out of 100. When the over-all gain is 13 db, the individual gains will not differ by more than about 0.4 db in 99 units out of 100. Probability studies involving each component in the converter were made in order that component tolerances might be as wide as possible and still permit the quoted tolerances in gain and therefore return loss to be met.

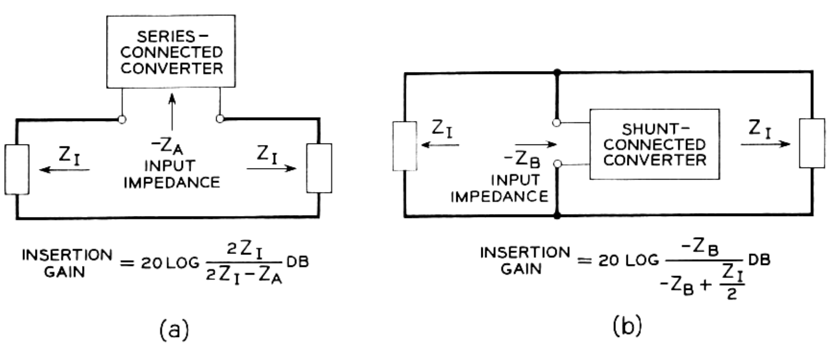


Fig. 34 — (a) Series-connected converter insertion gain; (b) shunt-connected converter insertion gain.

5.2 *Load Capacity and DC Stability*

The maximum output level that a repeater should be designed to carry is determined by cable crosstalk, allowable repeater gain and maximum talker volumes. From these points of view, it was found desirable that the E6 repeater carry approximately a +16 dbm single-frequency output level with relatively little compression.

In the E6 repeater the currents from the shunt and series converters add in phase in the receiving line and cancel in the sending line. Thus, all of the power from the converters is available at the receiving line. However, the maximum power available there decreases as the converter gain is increased. This is mainly because the gain-adjusting network of the series converter absorbs more power as the gain-adjusting network is increased to give more gain.

The maximum available power from the transistors is a function of the ambient temperature and their ability to dissipate heat due to the DC biases. The transistors used in the E6 repeater will stand 0.150 watt at 50°C and 0.110 watt at 60°C. Maximum ambient temperature is considered to be 50°C in most cases, but 60°C may be encountered in some unmanned desert offices. The DC biases are such that the DC power dissipation for each principal transistor is roughly 0.085 watt, so they are operated well below their rated output. The power dissipation in the piggy-back transistors is from 4 to 8 milliwatts.

The DC stability of the circuit is such that, when transistors with high leakage (i_{c0}) currents are used, the maximum power output is decreased about 0.5 db when the ambient temperature is 60°C. The compound transistor circuitry helps DC stability because most of the leakage current from the base of the principal transistor is returned to the collector by way of the collector of the piggy-back transistor without flowing through any bias resistors. Although the leakage current from the piggy-back transistor does disturb the biases, its effect is small because the DC power dissipation in this transistor is low.

5.3 *Susceptibility to Battery Noise*

The well-balanced push-pull circuits of the series and shunt converters of the E6 repeater permit the use of relatively noisy battery supplies. The noise susceptibility of a repeater may be expressed in decibels as

$$20 \log \left(\frac{\text{noise voltage across repeater output}}{\text{noise voltage across battery supply}}\right).$$

The noise in the output of the E6 repeater is at least 53 db lower than that in the battery supply at all frequencies between 60 to 3000 cps. Thus, a fairly noisy battery with a noise voltage of +45 db RN, as read on a Western Electric 2B noise set, would cause a repeater output noise of about (+45 − 53) db or −8 db RN.

5.4 *Lightning Surge Protection*

Lightning surge voltages are limited by the cable main frame protector blocks to about 600 volts peak. The average decay of surges is such that this peak voltage is decreased to about one-half in 600 microseconds. In some offices, depending on exposure and the number of storms, there may be as many as several hundred hits, with peak values from 200 to 600 volts, occurring in a single year. The total number of hits on a single cable pair during the lifetime of an E6 repeater may number in the thousands.

The circuits of the E6 converters are such that when 600-volt longitudinal or transverse surge is applied to its line terminals the maximum emitter-to-base (or collector-to-base) current through the transistors is about 0.5 ampere. The duration of this pulse is of the same order as that of the surge voltage (i.e., one-half peak value in 600 microseconds). The transistors used in the E6 repeater are designed to accept such surges without damage.

VI. APPARATUS DESCRIPTION OF THE E6 REPEATER

The mechanical design of the E6 repeater, a photograph of which is shown in Fig. 35, departs from conventional methods of packaging and is based on the matrix structure. In the E6 repeater, therefore, the printed wiring boards are used only for interconnection of the components into the desired electrical configuration as shown in Figs. 36 and 37. Molded phenolic blocks, sandwiched between the printed wiring boards, are used for mechanical support. Shaped cavities, within the molded blocks, are designed to accept the individual components. Two types of cavity are available within the blocks: one for pig-tail components and another for the larger components utilizing flexible leads.

The pig-tail cavities are so arranged that the axial dimensions of assembled component apparatus are at right angles to the printed wiring boards. To facilitate assembly, that portion of the block designed to accept pig-tail components is split into two equal and mating parts. The components are inserted into the lower half of the split block, which

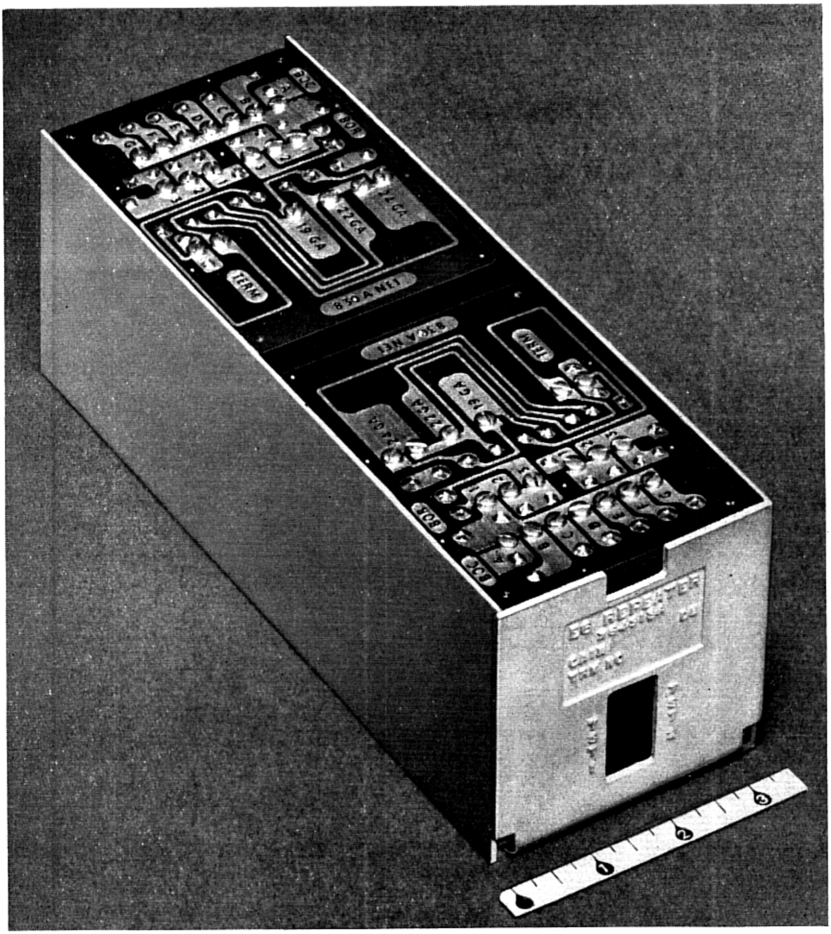


Fig. 35 — The E6 repeater.

acts as a carrier during manufacture. The top half is then dropped over the lower portion of the block in final assembly.

The portion of the phenolic block reserved for flexible lead components is not split, but is designed to supply open-top cavities into which components may be placed and subsequently potted. Connection to the printed wiring boards is made by means of feed-through terminals to which the flexible terminal leads of the components are soldered prior to encasement in a microcrystalline wax. In the converter the open-top cavities are protected by a metal plate, while in the LBO network this function is performed by one of the printed wiring boards.

To facilitate the insertion of pig-tail components, the individual cavities are tapered or funnel-shaped, with an additional tapered section at the base to guide the pig-tail lead wire through to the printed wiring board. The pig-tail leads are precut to proper length for connection to the printed conductors.

Transistors present the only exception to the cavity method of mounting, primarily because of the need for heat dissipation. A shelf was therefore provided on the lower block of the converter unit for external assembly of the transistors.

The converter unit of the E6 repeater, illustrating the features described above, is shown in Fig. 36. The lower portion of the phenolic mold is shown at the left of the photograph. The individual pig-tail components have been assembled in their individual cavities and the top portion of the pig-tail section is about to be placed in position. The open cavity section of the mold, reserved for flexible lead components, is also shown. In this photograph, the microcrystalline wax has not yet been poured into the open-cavity molded section. In normal production, however, this would have been done prior to the assembly of the pig-tail components.

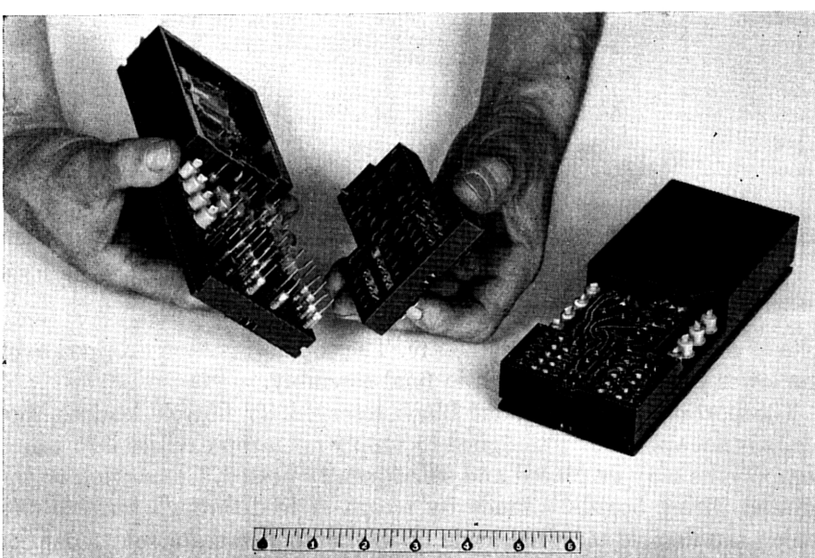


Fig. 36 — The E6 negative impedance converter.

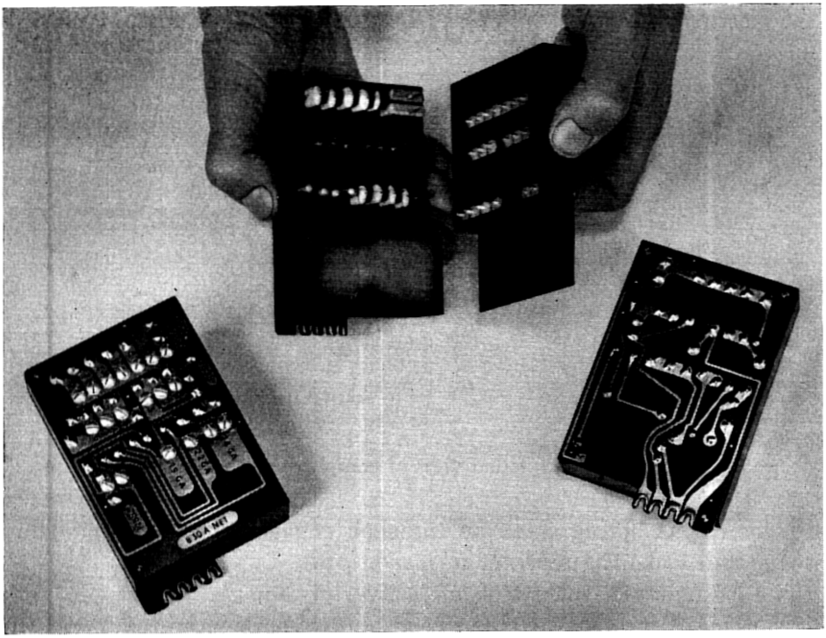


Fig. 37 — The 830A line building-out network.

Similar views of the LBO network are shown in Fig. 37. In this instance, the microcrystalline wax has already been poured and the unit is ready for assembly of the top section of the mold. In the other two views, the split-molded block is shown sandwiched between the top and bottom printed wiring boards of completely assembled LBO networks. The comb-like protrusion in the phenolic block, at one end of the LBO network, permits connection of the LBO networks to the converter. Electrical connection is made through open-ended brass grommets, soldered to the printed wiring board and arranged to accept the screws that go into tapped metal bushings located in the converter.

The E6 repeater is a plug-in unit, and all external connections are made by means of the plug and guidance details associated with the converter unit of the repeater, as shown at the right of Fig. 38. The mating connector is mounted on the equipment shelf. The connector assembly is such that a floating action at each terminal is obtained. When the repeater unit is plugged in on its mounting shelf, the tapered prongs of the guidance detail align the plug and the connector prior to

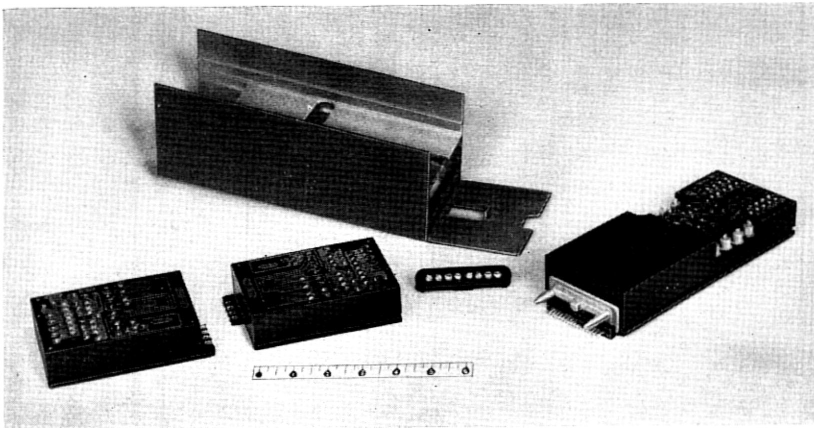


Fig. 38 — Exploded view of complete E6 repeater.

engagement. A spring catch on the shelf engages one of the guidance prongs and locks the repeater in place.

The E6 repeater consists of one converter unit and one or two LBO networks as dictated by systems use. As previously indicated, two LBO networks are required for intermediate use, while only one is required for terminal applications. The converter and LBO networks are housed in the aluminum H-beam extrusion of Fig. 38, which shows an exploded view of the complete E6 repeater. The bottom compartment of the H-beam extrusion accepts the converter unit, which is inserted guidance detail first from the front or hinged-door end. Grooves, molded in the sides of the converter unit, mesh with rails extruded along the inner surface of the H-beam. The transistors and adjusting screws, shown in the photograph are accessible at the bottom of the beam, when the converter is inserted in the housing.

The LBO networks are mounted in a similar fashion to that employed for the converter. The two networks, however, are each fed into the housing from opposite ends, one from the hinged-door end, and the other from the open end. The networks are positioned so that the comb-like protrusions of each network align in position above a filler block for screw connection to the converter.

For terminal applications, although only one LBO network is required, a dummy network must be supplied to provide a wiring path from one side of the converter to the line connections. In this instance, as shown in Fig. 39, the LBO network inserted from the hinged-door end is retained.

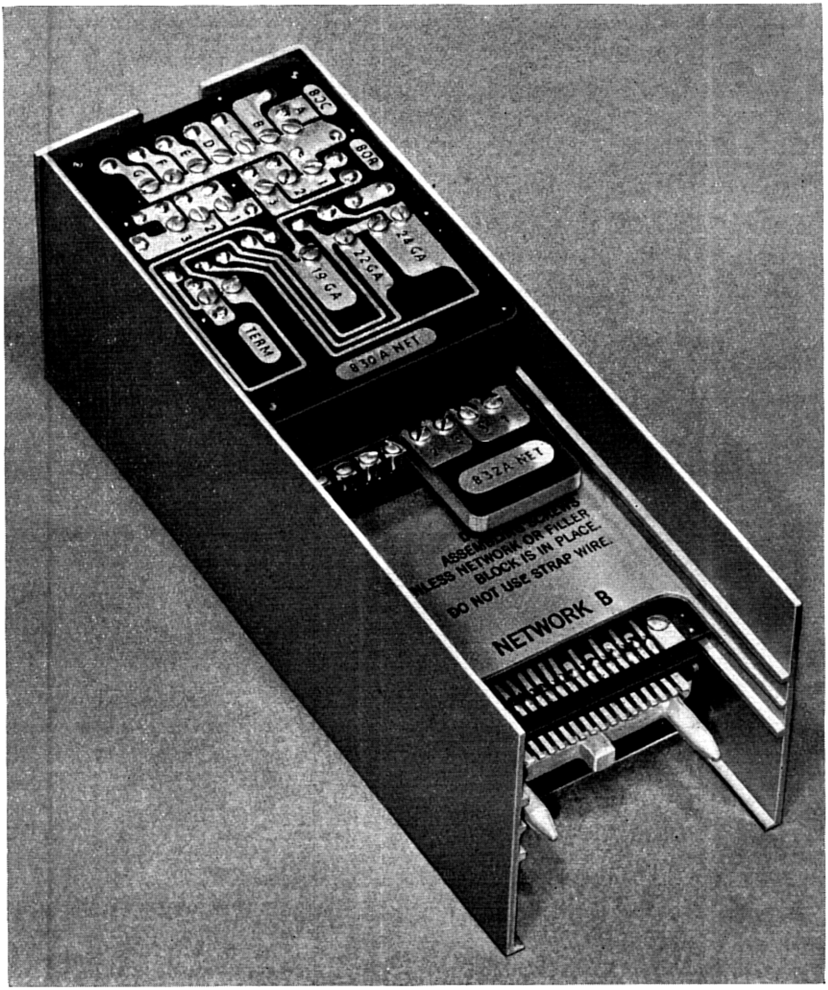


Fig. 39 — The E6 repeater — arranged for terminal use.

VII. TRANSMISSION TESTING EQUIPMENT

7.1 *Introduction*

One of the principal objectives of the E6 development was to reduce the cost of installing and maintaining negative impedance repeaters in the exchange area plant. In addition to the basic contributions of the E6 repeater toward this end, further simplification and economies have been realized by making available three new pieces of test equipment. These are the 54A transmission measuring set, the 54B test stand, and the 54C return loss measuring set.

7.2 *54A Transmission Measuring Set*

The 54A test set, shown in Fig. 40, is a portable, transistorized transmission measuring set used to determine the gain of the series and shunt converters of the E6 repeater — individually, or in combination. The importance of this function is evident when it is recalled that the attain-

Fig. 40 — 54A transmission measuring set.

ment of high return loss between the converter and LBO networks of the E6 repeater is dependent on the equality of gain settings of the series and shunt converters. The set may also be used for periodic measurements of gain and signal-level carrying capacity of the repeater.

7.3 *54B Test Stand*

The E6 repeater cannot be adjusted in position on the relay rack mounting shelf. In this position, the adjusting screws controlling converter gain and LBO network impedance-matching functions are not accessible. The 54B test stand, shown in Fig. 41, is therefore provided. The purpose of the stand is: (a) to hold the repeater in a convenient position for adjustment of the gain units and the LBO networks; (b) to supply DC power connections for the 54A transmission measuring set, the 54C return loss measuring set and the E6 repeater; and (c) to provide convenient connections and points of access for gain, return loss and trouble-locating tests.

The test stand is provided with a rotating platform into which the E6 repeater is inserted. The platform has two positions, each 180° removed from the other. In one position, the gain-adjusting screws of the converter are presented for adjustment, while in the other position the BOR, BOC and LFC network screws of the LBO network are made available.

Another feature of the test stand is the inclusion of a shunt holding circuit to prevent dialed-up test terminations from being released during test.

7.4 *54C Return Loss Measuring Set*

In order to attain the high degree of performance of which the E6 repeater is capable it is necessary that the LBO network and connected cable impedance match that of the E6 converter. In many applications, adequate return losses can be obtained at the LBO–converter junction by "book-value" adjustment of the BOR, BOC and LFC networks of the LBO network, based on office records of length of end-section and type of cable. This assumes, however, that the end-section value of the cable pair terminating a particular LBO network is accurately known. Experience on several field trials indicates, however, that the cable records are often not sufficiently accurate for this type of specification. The 54C return loss measuring set was, therefore, developed to permit optimum adjustment of the LBO networks on the basis of return loss measurements.

Fig. 41 — 54B test stand.

The 54C set, shown in Fig. 42, measures return loss in two frequency bands — either 500 to 2500 cps or 2000 to 3000 cps — with a self-contained sweep oscillator. Sweep frequencies are used because the relation between return loss and frequency can be highly variable and unpredictable in shape for specific cases. Such a sweep-frequency measurement

Fig. 42 — 54C return loss measuring set.

gives an average or integrated measurement that is more nearly representative than any arbitrary single-frequency measurement could be. The higher-frequency band is for BOC adjustment, which is more effective near cable cutoff; the lower and broader band for BOR and "gage" adjustments, which are effective across the whole band.

While the primary purpose of the 54C test set is for LBO network adjustment, it may ultimately be used in several general-purpose applications such as cable acceptance tests, adjustments of impedance compensators and precision networks associated with two-wire toll-connecting trunks.

VIII. ENGINEERING CONSIDERATIONS

In engineering trunk circuits for E6 repeater use, a number of factors must be considered if transmission loss objectives are to be met. Since the primary purpose of the E6 repeater is to provide voice-frequency amplification, methods of determining the maximum allowable converter unit gain consistent with singing, crosstalk signaling and overload requirements, etc. must be made available. The specification of the maximum allowable gain determines whether the circuit net loss requirements can be met, and will dictate, to a large extent, whether intermediate installation of repeaters is necessary. It then becomes necessary to determine if idle circuit terminations or repeater gain disablers are needed to prevent singing during idle periods. The effect of the E6 repeater on signaling must also be considered in the design of the circuit. Installation instruction must then be provided for an initial repeater adjustment so that a starting point for line-up procedures will be available.

Repeaters located at the terminal end of trunk circuits are limited to about 6.5 db net gain due to overload and crosstalk considerations. Those at intermediate locations may be operated up to 12 db net gain. To avoid singing in the idle circuit condition and to provide a satisfactory net loss, a computation of probable return loss at the converter terminals of the repeater is required. The return loss at the converter unit of the repeater is due to reflections at the central office end of the circuit and to reflections caused by cable irregularities. Estimates of return loss at the repeater require a detailed knowledge of cable pair loss, cable make-up and cable loading, since each irregularity is a point of reflection which degrades the return loss characteristic.

To evaluate the return loss at the converter terminals during the idle circuit condition, either of two approaches may be taken. In the first case, the far-end terminals are considered to be open circuit without idle circuit terminations. The terminal return loss is assumed to be 0 db and the maximum allowable gain is computed. This usually results in a circuit net loss greater than 3 db effective.* In the second case, where lower net losses are required, one practice is to terminate the circuit during the idle condition in a resistance at the far end. An alternative approach for this case is to provide a repeater disabler to reduce repeater gain when no supervisory current flows in the line, that is, when the line is idle and therefore unterminated. Under either condition, the repeater gain can be raised to the point where singing occurs in the interval between seizure of the circuit and the establishment of a through connection to the distant end. It is customary to assume 5 db return loss at a point equipped with an idle circuit termination. This usually permits raising the repeater gain by 1 or 2 db.

* Average of 15 selected frequencies throughout the voice band.

Having determined the maximum allowable gain, the circuit net loss must be computed. The net loss is the sum of the office, line and LBO network losses less the gain of the converter. The LBO loss, however, depends on the length and make-up of the end-sections of cable facing the repeater, which determine the amount of building-out capacitance and building-out resistance required in the LBO network.

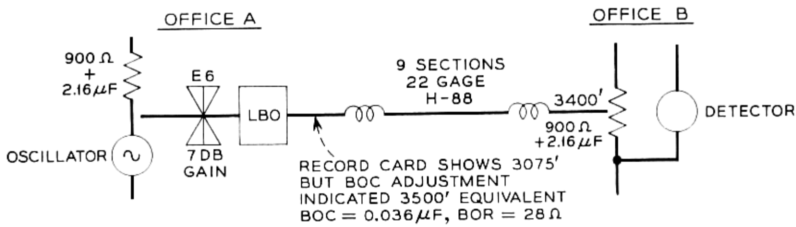
The effect of the E6 repeater on DC signaling must also be determined. The DC resistance of the converter and LBO networks introduces an adverse effect on signaling. The total shunt capacitance of the repeater introduces another. Some circuits, unfortunately, permit little margin for possible degradation of signaling due to repeater operation. In some instances, adjustment of signaling circuit relays may provide additional margins, but in others it may be found desirable to omit or reduce the BOR adjustment. This results in a degradation in return loss and thereby lowers the permissible gain.

Full realization of the return-loss capabilities of the E6 repeater can be obtained only if LBO network adjustments are made with a high degree of accuracy. The greatest problem encountered here is that of determining the value of end-section of a particular cable facility. Exact determination of cable end-section is complicated by office wiring and the mixture of cable gages that usually comprise the end-section. By use of the 54C return loss measuring set, however, it is possible to optimize the initial settings of the LBO network and to operate the converter at the maximum gain that cable irregularities will permit.

IX. SYSTEM RESULTS AND COMPARISON WITH E23 REPEATER

The E6 repeater will, in practically all cases, out-perform the older E23 repeater. With reasonable cable structural return losses, trunks equipped with E6 repeaters can be made to meet the systems requirements for over-all net loss and for return loss to the toll office balancing network. Much of the improved performance of the E6 repeater, as compared to the E23 repeater, is due to the range and precision of adjustment available in the LBO networks. Another contributing factor to improved performance is that the image impedance of the converter approximates 900 ohms + 2.16 microfarads and is essentially constant with gain setting.

Trunks equipped with E6 repeaters can usually be operated at 1 to 2 db lower net loss than can trunks with E23 repeaters. This lower loss is obtainable with equal or improved return losses at the ends of the trunk. In addition to the better performance, it is possible, by use of careful adjustment procedures, to raise the gain of E6 repeaters by a



MEASURED NET LOSS

	1000 CYCLE		EFFECTIVE	
	WITHOUT REPEATER	WITH REPEATER	WITHOUT REPEATER	WITH REPEATER
LAB SETUP	8.3 DB	1.7 DB	8.6 DB	2.3 DB
ACTUAL	9.0 DB	2.3 DB	9.2 DB	2.9 DB

RETURN LOSS AT OFFICE A WITH COMPENSATOR NETWORK TERMINATION AT OFFICE B

	500–2500 ∿	2000–3000 ∿	MINIMUM	CABLE STRUCTURAL RETURN LOSS
ACTUAL	20.3 DB	14.4 DB	8.3 DB AT 2900 ∿	38.1 DB
LAB SETUP	20.6 DB	19.0 DB	17.0 DB AT 2900 ∿	> 40.0 DB

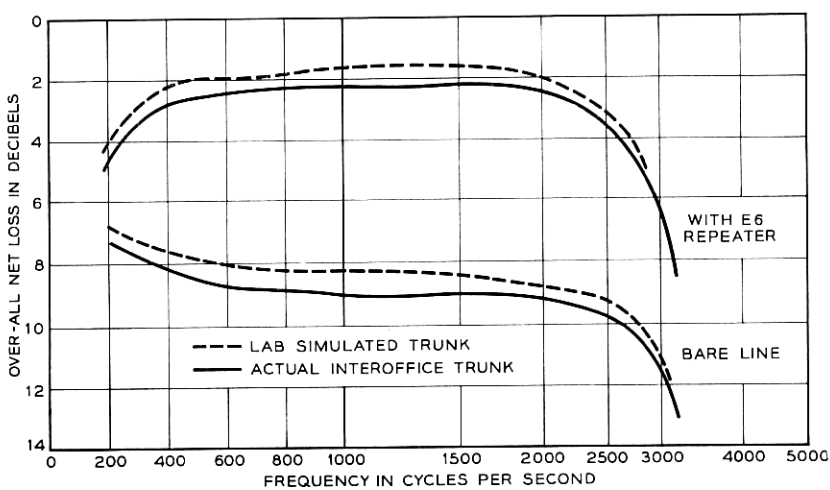
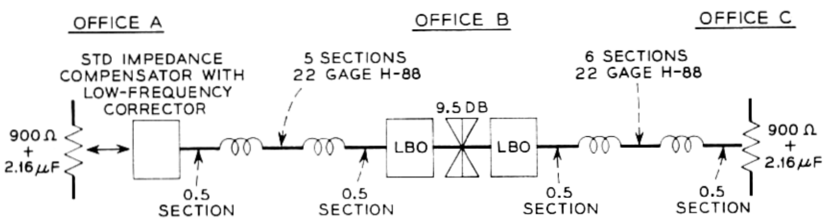


Fig. 43 — Performance of E6 repeater terminal on nine sections of 22-gage H-88 cable, frame to frame.

sufficient amount so as to remove most of the flat-with-frequency office losses at each end of the trunk.

A number of illustrations are given to show the performance of the E6 repeater. Fig. 43 is a comparison of test results obtained on a terminal repeater using physical and artificial cable. The repeaters were adjusted in such fashion as to maintain stability under all combinations of passive terminations at the trunk terminals. Where the structurals of the cable



MEASURED NET LOSS OFFICE A TO OFFICE C

1000 CYCLE		EFFECTIVE	
WITHOUT REPEATER	WITH REPEATER	WITHOUT REPEATER	WITH REPEATER
9.8 DB	1.5 DB	10.1 DB	2.5 DB

RETURN LOSS AT OFFICE A WITH COMPENSATOR NETWORK TERMINATION AT OFFICE C

500–2500 ∿	2000–3000 ∿	MINIMUM
19.5 DB	19.5 DB	14.5 DB AT 2500 ∿
WITH PRECISION NETWORK AT OFFICE C		
23.0 DB	23.0 DB	——

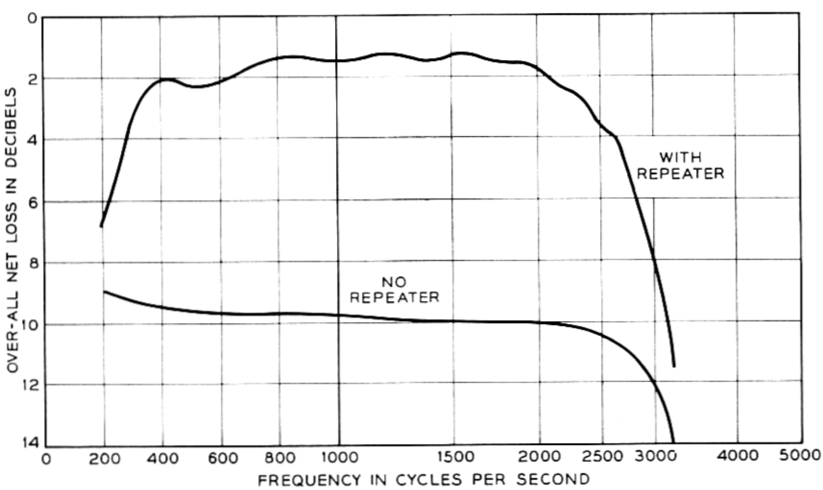
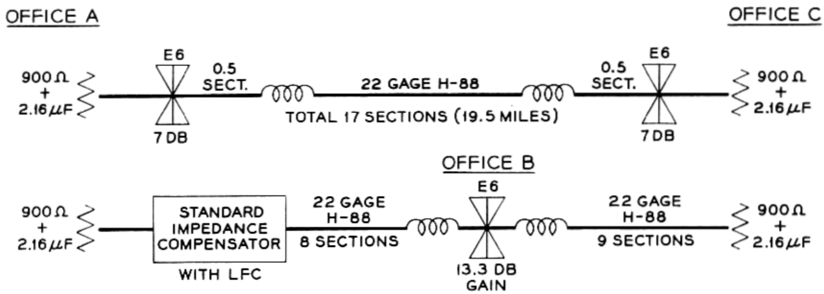


Fig. 44 — Performance of E6 repeater intermediate on 22-gage H-88 cable with impedance compensator.

were high, the results for both conditions are reasonably close and quite satisfactory.

On longer trunks requiring more than 6.5 db gain, an intermediate repeater with two LBO networks is required. Results on simulated cable are shown on Fig. 44. In this case a standard impedance compensator is used at the toll or office A end to improve the return loss.



RETURN LOSS AT OFFICE A

SWEEP	0.5 TO 2.5 KC	2-3 KC
2 REPEATERS	21.2 DB	21.8 DB
1 REPEATER	19.5 DB	20.5 DB

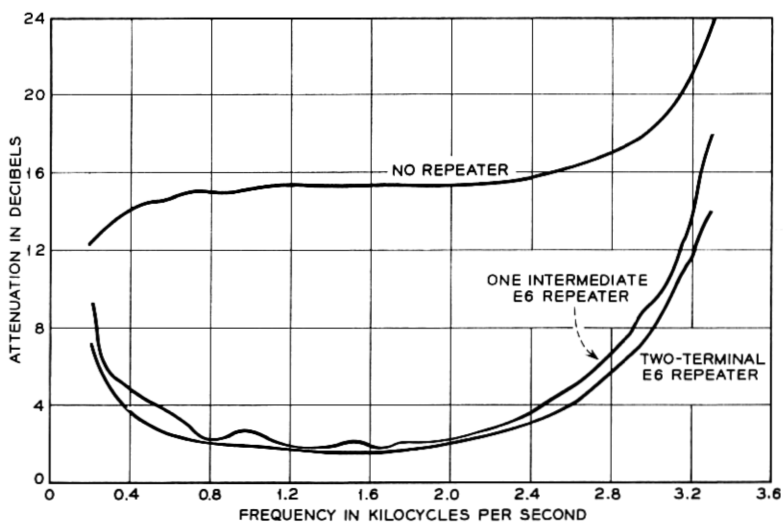


Fig. 45 — Over-all response with E6 repeaters at each end compared to a single intermediate E6 repeater.

For longer trunks with losses greater than 15 db, the best results are obtained with a terminal repeater at each end of the circuit. An intermediate repeater, giving somewhat inferior results to the two-repeater installation, may also be used if economies are desired. Results obtained for intermediate and two-terminal repeater operation, using simulated cable, are compared in Fig. 45. Return losses in either case are satisfactory when high cable structurals are available.

In the following examples, direct comparisons are made between E6 and E23 repeaters operating under field conditions. The first example is that of a tandem-completing trunk utilizing one repeater at an intermediate point. The circuit layout and the over-all frequency response, including effects of office equipment, are shown in Fig. 46. Each repeater

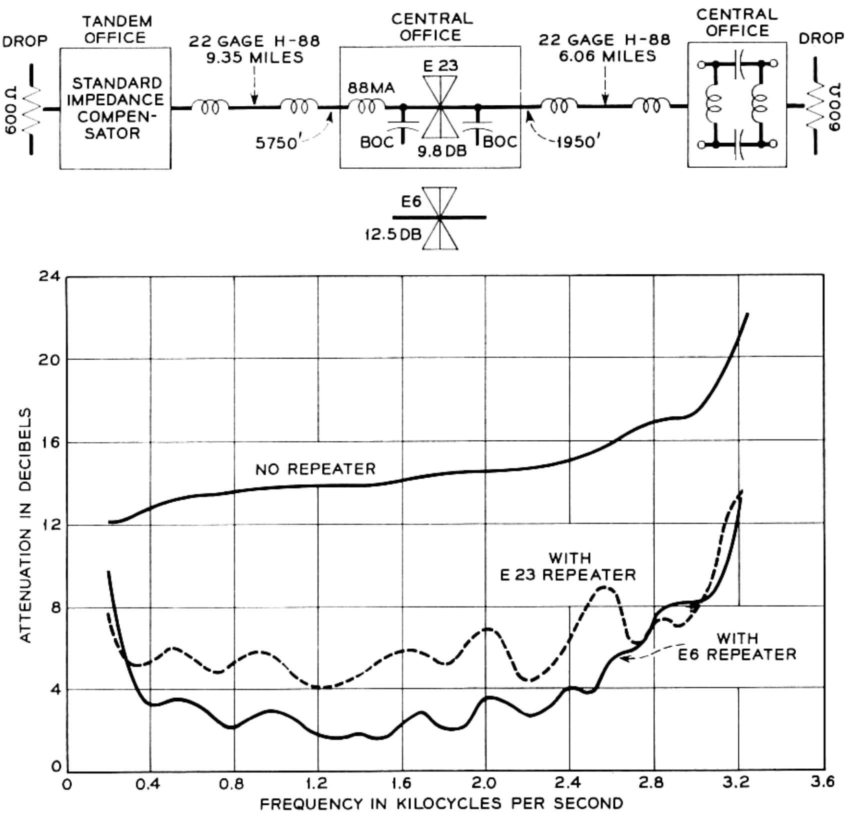


Fig. 46 — Over-all response including office equipment of tandem-completing trunk with intermediate E6 repeater.

was adjusted for maximum stable gain, using both idle and talking circuit terminations on the drop side of each office. Most of the office losses, amounting to approximately 1.3 db, were therefore removed. Stability of the repeater was also checked during the switching period to insure that singing did not occur during the interval required to establish a connection. The improvement of the E6 over the E23 is evident from the characteristics of Fig. 46. Measurements were made using 600-ohm apparatus and, as a result, certain irregularities in the response are to be observed. Had 900-ohm apparatus been used, lower net losses and smoother curves would have resulted due to the elimination of reflections. The E23 response is only as good as shown because a loading coil and a building-out capacitance equivalent to 0.5 end-section were added to the line on the 5,750-foot end-section side of the repeater. Without this circuit modification, a characteristic which falls off very rapidly at

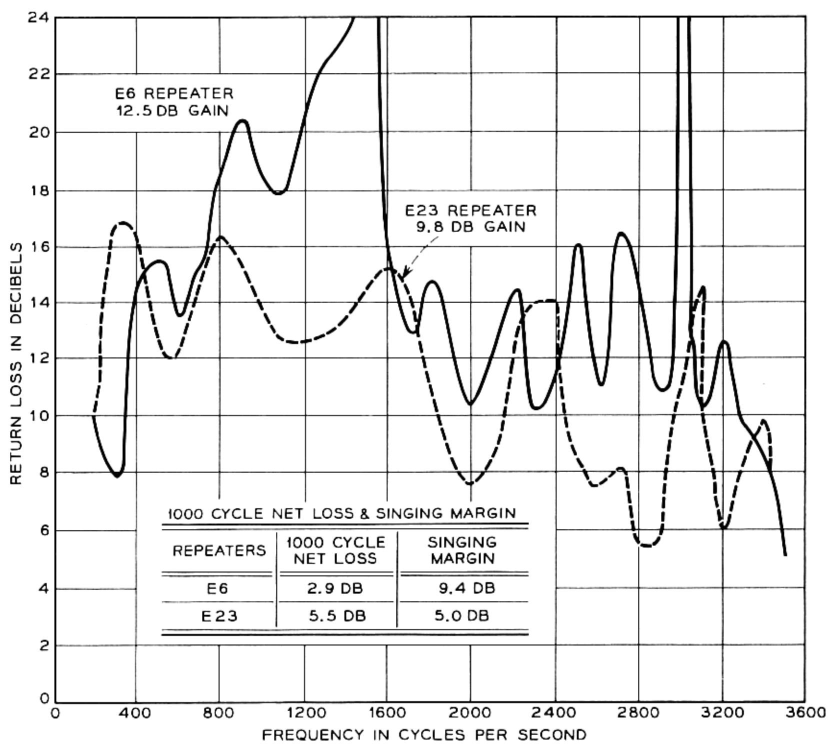
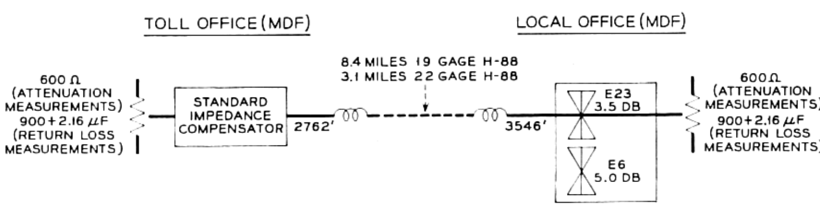


Fig. 47 — Return-loss characteristics of E6 and E23 repeaters as measured in field test.

the high end would have resulted. On the other side of the repeater some building-out capacitance was also required to bring the end-section up to 0.5-section. When the E6 repeater was used, however, no loading coil or external building-out capacitance was necessary, and each LBO was adjusted to full end-section with the elements available in the network.

In addition to the lower net losses made possible with the E6 repeater, better return losses and more adequate singing margins are also obtained. A plot of the return-loss characteristic taken under the field test conditions described above is shown in Fig. 47. The improved return loss response with the E6 repeater is apparent.

An example of a toll-connecting trunk terminal application is shown in Fig. 48. The repeater is located at the local or class 5 office. In this case, the trunk is made up of a mixture of 19- and 22-gage cable. Again, the E6 repeater gives a lower circuit net loss, amounting to an improvement of approximately 2 db over the E23 repeater. In spite of the differ-



RETURN LOSS (TOLL OFFICE)

REPEATER	FREQUENCY RANGE		
	0.5 - 2.5 KC	2-3 KC	S.P. (DROP-TO-DROP)
E6	18.2 DB	15.7 DB	12 DB
E 23	18.5 DB	15.7 DB	11 DB

RETURN LOSS (LOCAL OFFICE)

REPEATER	FREQUENCY RANGE	
	0.5 - 2.5 KC	2-3 KC
E6	19.0 DB	16.0 DB

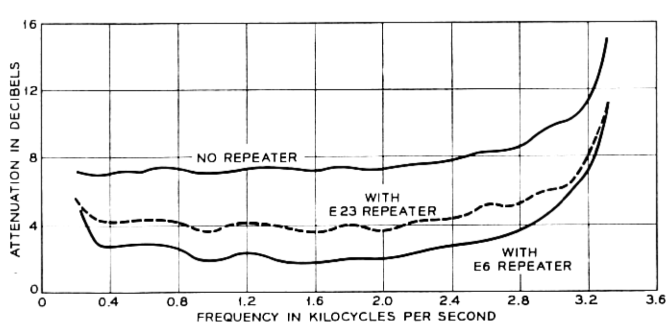


Fig. 48 — Over-all response of a toll-connecting trunk with terminal repeater at a local office.

ence in net loss, both repeaters gave about the same return loss results at the toll office end. At the repeater end, however, the E6 repeater gave considerably better return losses.

X. ACKNOWLEDGMENTS

The authors wish to acknowledge the many contributions made by their associates at Bell Telephone Laboratories, Western Electric Company and the American Telephone and Telegraph Company in the development of the E6 repeater. In particular, they wish to acknowledge the contributions of R. W. DeMonte to LBO network design techniques, W. C. Schmidt to packaging concepts and packaging development and J. K. Jones to equipment design and layout.

REFERENCES

1. Merrill, J. L., Jr., Rose, A. F. and Smethurst, J. O., Negative Impedance Telephone Repeaters, B.S.T.J., **33,** 1954, p. 1055.
2. Merrill, J. L., Jr., Theory of the Negative Impedance Converter, B.S.T.J., **30,** 1951, p. 88.
3. Gerardi, B. and Jewett, F. B., Telephone Repeaters, Proc. A.I.E.E., **38,** 1919, p. 1255.
4. Crisson, G., Negative Impedance and the Twin 21-Type Repeater, B.S.T.J., **10,** 1931, p. 485.
5. Linvill, J. G., Transistor Negative Impedance Converters, Proc. I.R.E., **41,** 1953, p. 725.
6. Darlington, S., U. S. Patent No. 2,663,806.
7. Fano, R. M., Theoretical Limitations on the Broadband Matching of an Arbitrary Impedance, J. Frank. Inst., **249,** 1950, pp. 57; 139.
8. Matthaei, G. L., Synthesis of Tchebycheff Impedance Matching Networks, Filters and Interstages, I.R.E. Trans., **CT-3,** 1956, p. 163.
9. Kuh, E. S., unpublished memorandum.
10. DeMonte, R. W., unpublished memorandum.
11. Guillemin, E. A., *Communication Networks*, Vol. 2, John Wiley & Sons, New York, 1935, p. 356.