# E7 REPEATER
# THEORY

## 1. LOOP CHARACTERISTICS

**1.01** Loops consisting of nonloaded cable between central office and subscribers take many forms. They may be simple, consisting entirely of a single gauge of conductor, or they may be complicated, containing two, three, or four different gauges. Many contain bridged taps for multiparty service and for flexibility in assignment.

**1.02** When terminated in a station set at the customer's premises, however, the central-office-end admittance of any loop, simple or complicated, exhibits a frequency characteristic typical, in shape, of all nonloaded loops. The loss and the phase shift between the CO and the subscriber also exhibit the same characteristic shape, which is illustrated in Fig. 1.

**1.03** The changes in the CO-end input admittance, $Y_c$, loss, and phase shift, between about 0.3 and 3.0 kHz are the causes of undesirable distortion in transmission, and of undesirably low return loss of loop vs. CO. If each of these characteristics remained essentially constant throughout that frequency range, the loop's admittance could be matched to that of the central office by inserting a transformer of proper ratio at the office end of the loop. Also, the susceptance and phase shift ("b" and "$\beta$" in Fig. 1) would become trivial. But, since the admittance, the loss and the phase shift do not remain constant, it is necessary to do much more than insert a simple transformer.

## 2. THE IDEALIZED REPEATER

**2.01** Although $Y_c$, the CO-end input admittance of the loop, is usually a complicated function of frequency, it can be approximated fairly well in the voice range by a simple 3-element, 2-terminal network shown in Fig. 2. The resistance $R_1$ represents the desired ideal, while the RC branch contributes the unwanted frequency distortion.

**2.02** If another, negative, RC branch of equal magnitude were inserted in parallel with the RC branch, as shown in Fig. 3, the two together would present infinite impedance across terminals 2-2. $Y_c$ would then consist of $R_1$ only. A suitable transformer would then match $R_1$ to $R_o$, the ***ideal*** office impedance. If an ideal negative-impedance converter (NIC) and an ideal transformer were available, as shown in Fig. 4, the gain attainable through use of the negative RC branch would be given by the equation

$$G(\text{db}) = 20 \log \frac{N}{2} (1 + Z_oY_c)$$

where $Z_o$ is the ***actual*** office impedance and N is the turns ratio of the transformer. If $Z_o = \frac{1}{2Y_c}$ and $N = 2$, the gain is 3.5 db. The equation is general and does not depend on the configuration of the negative impedance. In a real application, N and $Z_o$ are constant, while $Y_c$ increases with frequency. Thus, the gain increases with frequency, as required for equalizing transmission.

### Inductive Coupling of NIC

**2.03** The NIC might be coupled to the line by means of a transformer and a capacitor to block direct current as in Fig. 5. As good or better performance, however, can be had by inductively coupling it to the transformer in Fig. 3 through the use of a third winding as in Fig. 6, thereby avoiding the need for the second transformer and a blocking capacitor.

### Matching the Office Impedance

**2.04** Fig. 6 indicates that $R_1$ is being matched to $R_o$, the ideal office impedance, but in practice $R_1$ should be matched to $Z_o$, the actual office impedance. The transformer's office and line windings also require means for blocking direct current. Both requirements are met by the use of a 1-$\mu$F capacitor. The two windings are split in the center and half windings are connected in series as shown in Fig. 7. The capacitor is connected between the midpoints to prevent a dc shunt connection between the two wires of the loop.

### Equivalent Dissymmetrical Bridged-T Structure

**2.05** In Fig. 8 the circuit is drawn in the unbalanced form. For purposes of analysis, it can be further simplified by assuming the NIC to be ideal, that is, equal to a multiplier of −1, and by bridging the negative impedance directly across the dissymmetrical T configuration shown in Fig. 9. The mutual inductance M in series with $C_s$ is negative, hence the total shunt reactance is always negative. In general, transformers are connected across the line in such a manner that they introduce much loss at low frequencies. In this application, relatively high reactance at low frequencies comes from the capacitor $C_s$ and, therefore, the transformer is permitted to operate with smaller mutual inductance. At frequencies above 1 kHz, the reactance of $C_s$ is small compared to that of the mutual M and hence its effect is small. The gain at very low frequencies is inserted in series because $C_s$ introduces a high reactance compared to −M. Also, $C_o$ introduces a large reactance and $C_s$ aids substantially in producing an impedance match. At high frequencies, $C_s$ can be assumed shorted so that practically all the negative impedance is across the line in the manner indicated in Fig. 3.

### Principle of the Negative-Impedance Converter

**2.06** The active negative-impedance converter (NIC) is a gain unit connected in the circuit by means of inductive coupling to the transformer. This converter is of the short-circuit-stable type, meaning that if the output terminals of the converter are short circuited, the device will not oscillate or sing. It has long been known that it is possible to obtain circuits with this property of converting positive to negative impedances by coupling the output of an amplifier back to its input. Consider a special kind of balanced amplifier having a current-amplification ratio of K and a negligibly low input impedance. If this amplifier is connected as shown in Fig. 10, a short-circuit-stable negative impedance is seen toward the right from Terminals A and B, as shown in the following development.

$$\begin{aligned} E &= Z_o(K-1)i - \frac{Z_N}{2}i - \frac{Z_N}{2}i \\ &= Z_o(K-1)i - Z_N i \end{aligned}$$

Also, $E = Z_o(K-1)i + Z_{AB}(K-1)i$

Therefore $Z_o(K-1)i - Z_N i = Z_o(K-1)i + Z_{AB}(K-1)i,$

$$Z_{AB}(K-1)i = -Z_N i,$$

$$\text{and } Z_{AB} = -\frac{Z_N}{K-1}$$

If Terminals A and B are short-circuited, no voltage can exist between those terminals or across the amplifier's input terminals, hence the device is short-circuit stable.

## 3. THE E7 REPEATER

### Push-Pull Transistorized Negative-Impedance Converter

**3.01** The input and output impedances of a push-pull type common base transistor-amplifier closely approximate the balanced amplifier just described; the input impedance is low, the output impedance very high, and the current ratio between emitter and collector very nearly equal to 1. The circuit in Fig. 10 then, is equivalent to that shown in Fig. 11. If N is the turns ratio of the transformer, then the voltage across AB will equal $iZ_N$ and the voltage across $CD = \frac{iZ_N}{N}$. The impedance at

$$CD = -\frac{\frac{iZ_N}{N}}{iN} = -\frac{Z_N}{N^2}$$

Note that a short circuit across CD means zero voltage across winding EF.

**3.02** Push-pull type circuits provide the feedback paths of the proper phase and cancel even-harmonic distortion products. The capacitors $C_1$ and $C_2$ couple the converter to the transformer and provide the feedback connection. Their capacitance is sufficiently large to have negligible impedance over most of the voice-frequency range. The mutually coupled inductor $L_2$ is used for shaping the impedance at high frequencies. The converter cuts off at about 20 kHz. This means that, as frequency increases, the conversion ratio is no longer −1 but gradually decreases numerically and becomes a complex number. The real part of the number eventually changes sign from negative to positive and the converter with its terminating impedance $Z_N$ behaves like an ordinary series RC impedance. The center-tapped winding acts as a balancing inductor to maintain stability by providing a low impedance at this point over a wide frequency band. The center tap also provides means of supplying the dc potential to the transistors.

### Inductive Coupling to Cable Pair

**3.03** When a third winding is added to the transformer in Fig. 11 and that winding is connected to the cable pair as shown in Fig. 7, two practical objectives are accomplished. One is the removal of dc supervisory voltage from the negative-impedance circuit and the other is the ability to transmit pulses through the transformer. The transformer coupling gives a high degree of isolation of the converter from the effects of power-induced longitudinal voltages, which in a small percentage of lines may reach values measured in tens of volts. Lightning-induced surges may reach values of 350 to 500 volts before the carbon protector blocks break down. To minimize the effects of such surges, transformer coupling is valuable.

### Use of Compound Transistors

**3.04** In the preliminary analysis of the transistor-converter it was assumed that all signal current flowed through the emitter and collector and none through the base circuit. Actually the emitter and collector currents can differ by a few per cent. The negative impedance viewed from the line terminals is more accurately given by $Z_{input} = (1 - 2\alpha)Z_N + (1 - \alpha)2r_b$ where $r_b$ is the base resistance of the transistor and the impedances of $C_1$ and $C_2$ are small. The $\alpha$ of good transistors varies between 0.95 and 1 so that the conversion factor is dependent on the value of transistor $\alpha$. This may be of small consequence where each individual repeater circuit is adjusted separately, but where a closely reproducible negative impedance is desired for a given value of the network resistance, close limits are required on the $\alpha$ characteristic of the transistor and its variation with time. In some typical networks for loaded cable use, a change in $\alpha$ from 1 to 0.95 would change the ratio of $\frac{Z_{input}}{Z_N}$ from 1 to 0.878. Such variations can be markedly reduced by a tandem or double-transistor arrangement shown in Fig. 12 which returns a very large percentage of the base current back to the emitter-collector circuit of transistor 1.

**3.05** The arrangement is as follows: Each transistor is replaced by a compound transistor in which the base current of the first transistor is returned through the emitter-collector circuit of the second transistor to the desired path. The losses in base current are now a function of $(1 - \alpha_1)(1 - \alpha_2)$ which is much smaller than the $(1 - \alpha)$ obtained with a single transistor. As a result, the change in conversion factor, $\frac{Z_{input}}{Z_N}$, for a variation in transistor $\alpha$ from 1 to 0.95, is only from 1.000 to 0.995. With a single transistor this much variation in $\alpha$ would result in a change from 1 to 0.878. For the compound circuit, an $\alpha$ of 0.9 for the transistors results in a conversion ratio of 0.987. Thus, by using twice as many transistors, it is possible to widen the tolerance for $\alpha$ and to make drastic reduction in the effects of variation of the $\alpha$'s as the transistors age. The wider tolerance makes it practicable to use standard transistors instead of more expensive ones especially selected for $\alpha$'s close to 1.

### Leakage Inductances

**3.06** As is well known, all practical transformers have a certain series impedance as a result of leakage inductance and heat losses in windings. In some applications this impedance is not objectionable and may even be helpful, as in the case of open-circuit-stable negative-impedance converters. In the case of short-circuit-stable negative-impedance repeaters, however, the reactive component may cause circuit instability, particularly at frequencies above the voice range. One method of overcoming this effect is to introduce a negative impedance equivalent to this series impedance. As shown in Fig. 13, $R_e$ and $L_e$ are the resistance and inductance components referred to the winding connected to the NIC.

**3.07** Assuming an NIC with a multiplier of $-1$, elements of value $R_e$ and $L_e$ can be connected in series with the terminating network of R and C to cancel the undesirable leakage effect and resistance of windings. Physically, $R_e$ is combined with R.

### Overall Stability

**3.08** In practice, the repeater must be capable of operating stably for a wide variety of cable facilities having a wide variety of impedance characteristics. Selection of the optimum transformer tap and the values of R and C in the terminating network may produce instability if the gain characteristic does not have the proper decreasing slope at high frequencies. One method of avoiding instability is to modify the terminating network. A more economical and simpler method is to add a resistance of 30 ohms in series with the 1-$\mu$F capacitor. This resistance not only shapes the gain curve, but also improves the return-loss characteristics at high frequencies without appreciably affecting the characteristics in the pass band.

### Stability Margin

**3.09** When lined up on a specific circuit, the repeater must remain stable, whether idle or in service, and regardless of the temperature of the outside facilities. The idle-circuit terminations are often quite different from the in-service terminations, and the impedance characteristics are usually quite different under the two sets of conditions. Seasonal temperature variations are appreciable and cause important variations in propagation and impedance of the cable facility. Obviously, if a repeater is lined up in hot weather, its insertion-loss and return-loss performance in cold weather will not be as good. Engineering consideration was given to these variations as well as the unavoidable manufacturing variations among repeaters, in order to provide a reasonable margin of stability.

### Complete Schematic of Repeater

**3.10** A complete schematic is shown in attached Drawing SD-99705-01. The magnitudes of the base resistors R3, R4, R5, and R6 are set from considerations of compound transistor action, load-carrying capacity, terminating impedance, and reduction of peak currents during lightning surges. Resistors R1 and R2 divide the 48-volt battery for the proper voltage biases on all the transistors. Resistors R7 and R8 are also part of the biasing network. Since their resistance is large compared to the impedance of the adjustable network (R10 and C4-C10), they have only a slight shunting effect on it.

## 4. DESCRIPTION OF E7 REPEATER

**4.01** (a) The E7 is a transistorized, 2 wire, voice-frequency telephone repeater of the plug-in type.

(b) It is designed to be inserted between the central office and a subscriber's non-loaded line.

(c) Its purpose is to provide transmission-frequency equalization and to improve the return loss at the office. It was not designed primarily to give gain at 1 kHz, and generally gives less than 1 db at that frequency.

(d) Its impedance viewed from the central office is capable of adjustment to give about 20 db echo return loss against 900 ohms in series with a 2-$\mu$F capacitor.

(e) The net transmission loss is primarily determined by selection of the turns ratio of the coupling transformer which corresponds approximately to the following gains at 2300 kHz.

| TAP | TURNS RATIO | GAIN IN DB |
|---|---|---|
| 1 | 1.1:1 | .84 |
| 2 | 1.41:1 | 3.00 |
| 3 | 2.0:1 | 6.00 |
| 4 | 2.5:1 | 7.96 |

**4.02** For a given turns ratio, the return loss is primarily determined by the value of the elements in the terminating network. These values are selected by means of a number of screw-type switches. The range of resistance is from 30 to 180 ohms, smoothly adjustable. The range of capacitance is from 0 to 1.6 $\mu$F in steps of approximately 0.01 $\mu$F.

**4.03** Operation of the repeater requires battery supply of −48 volts. Ambient temperatures should not exceed 140° F. DC pulse signaling currents range up to .15 amp.

**4.04** Two MON jacks, accessible from the front of the repeater, provide for monitoring with a specific high-impedance headset.

**4.05** Three test points, TP1, TP2, TP3, accessible only with the repeater removed from the shelf, provide for transistor bias tests.

**4.06** The cover for the E7 repeater is an extruded H-shaped shell with a die-cast front door hinged at its lower edge. The shell and door are of aluminum alloy in natural color. Rails along the inside of the cover sidewalls hold the two blue epoxy boards, the top one being the terminating network, and the bottom one the converter, including the coupling transformer. The boards are held in place by two machine screws, which also provide the electrical connections between the two boards. The bottom edges of the cover sidewalls are turned inward to fit grooves formed by guides on the repeater shelf.

**4.07** An opening in the door provides access to pinjack test points. A notch in the top edge of the door provides for insertion of a finger tip to open it. Connections to power supply, CO, and loop facilities terminate in a plug at the back of the repeater. Rugged pins near the plug guide it into proper engagement with a connector on the repeater mounting shelf, relieve it and the connector of possible strain due to repeater weight, and engage a spring catch on the shelf to lock the repeater in place. The plug-in feature facilitates testing and maintenance, particularly where test jacks are not provided, as it allows easy removal of a defective repeater to a location more convenient for maintenance, and replacement with a spare repeater. The door design eliminates the need for a tool to free the repeater from the spring catch. When the door is opened, a cam along its lower edge pushes against the front edge of the guide on the shelf. This moves the repeater outward a sufficient distance to unlock it from the spring catch and free it from the shelf connector.

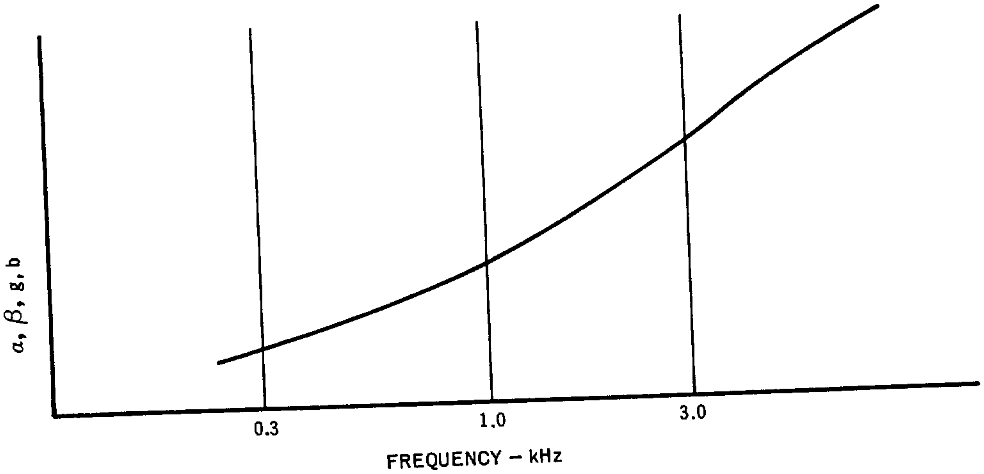


Fig. 1

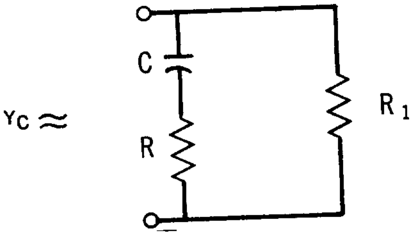


Fig. 2

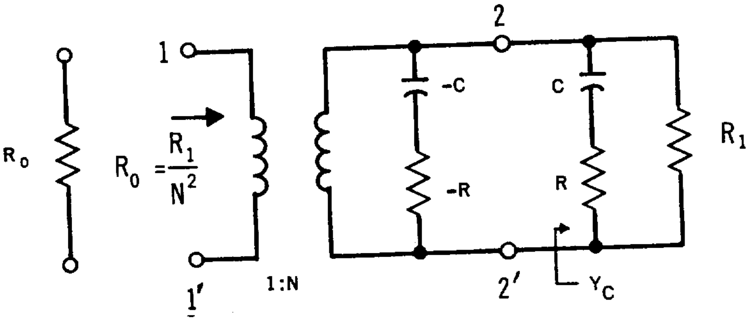


Fig. 3

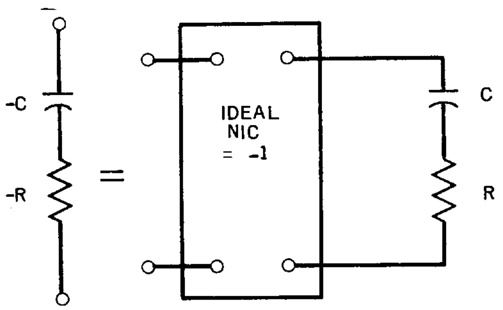


Fig. 4

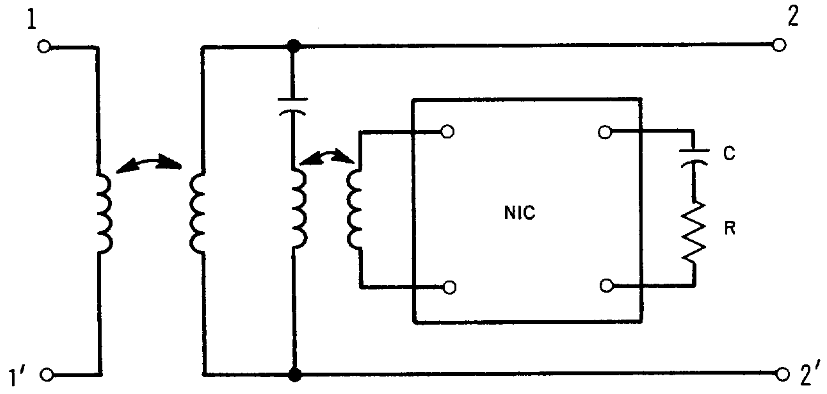


Fig. 5

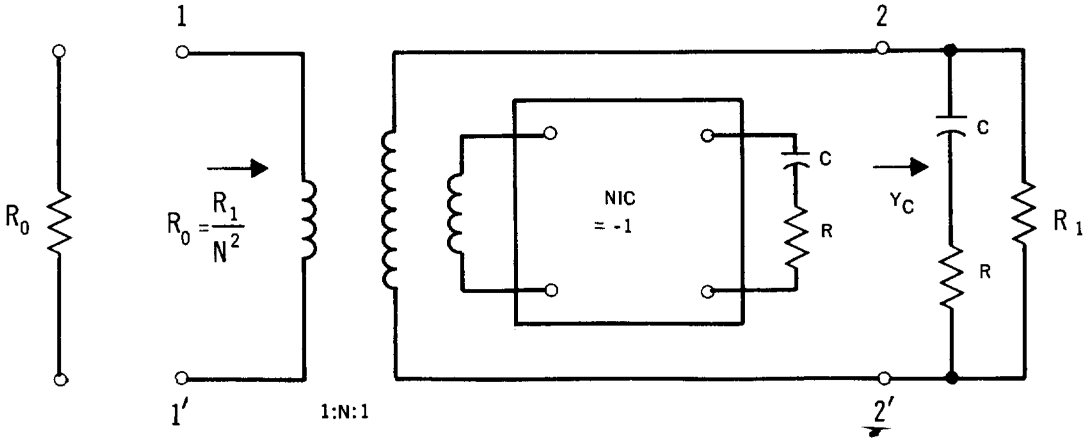


Fig. 6

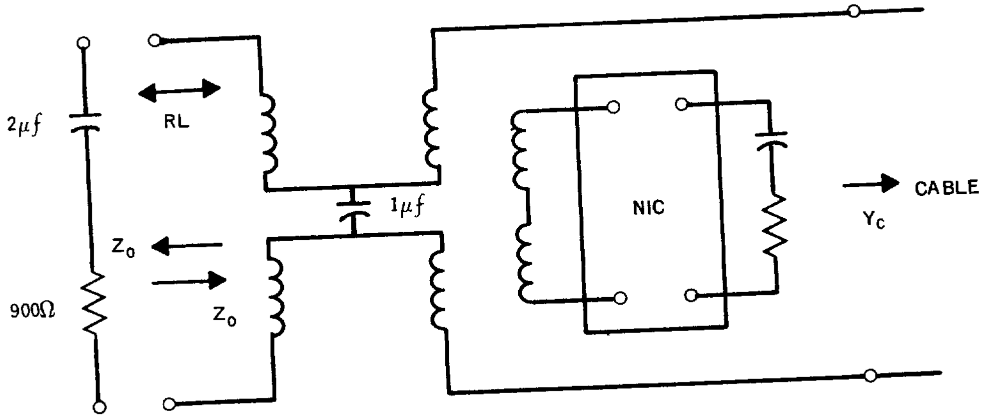


Fig. 7

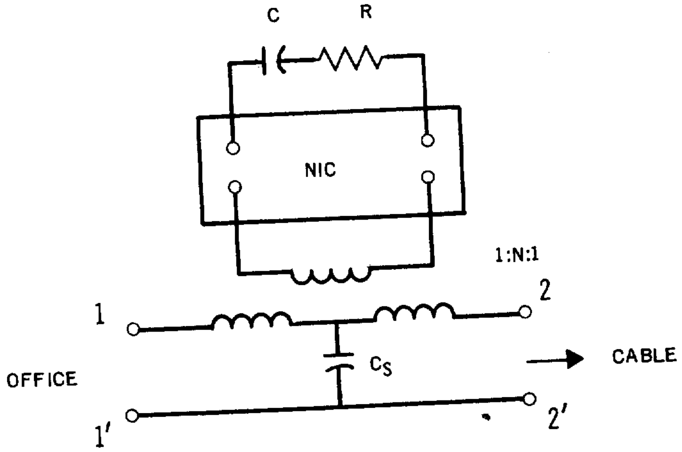
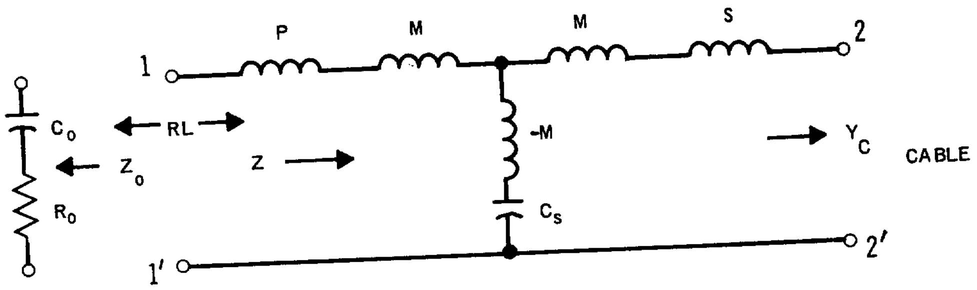
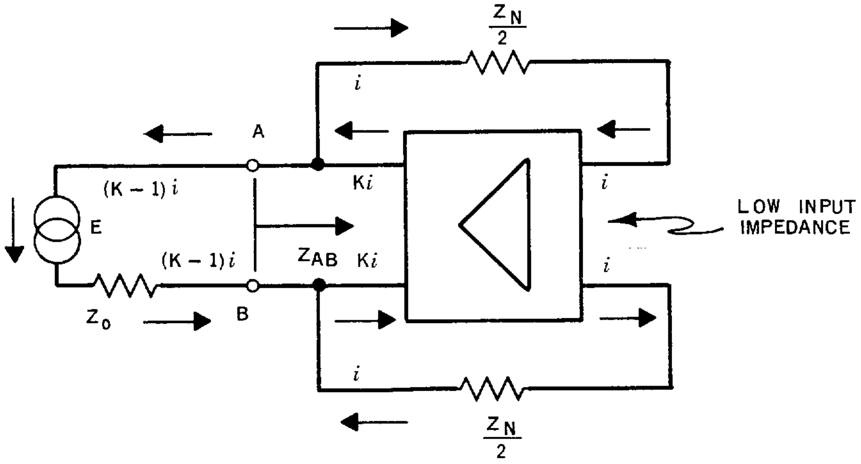


Fig. 8



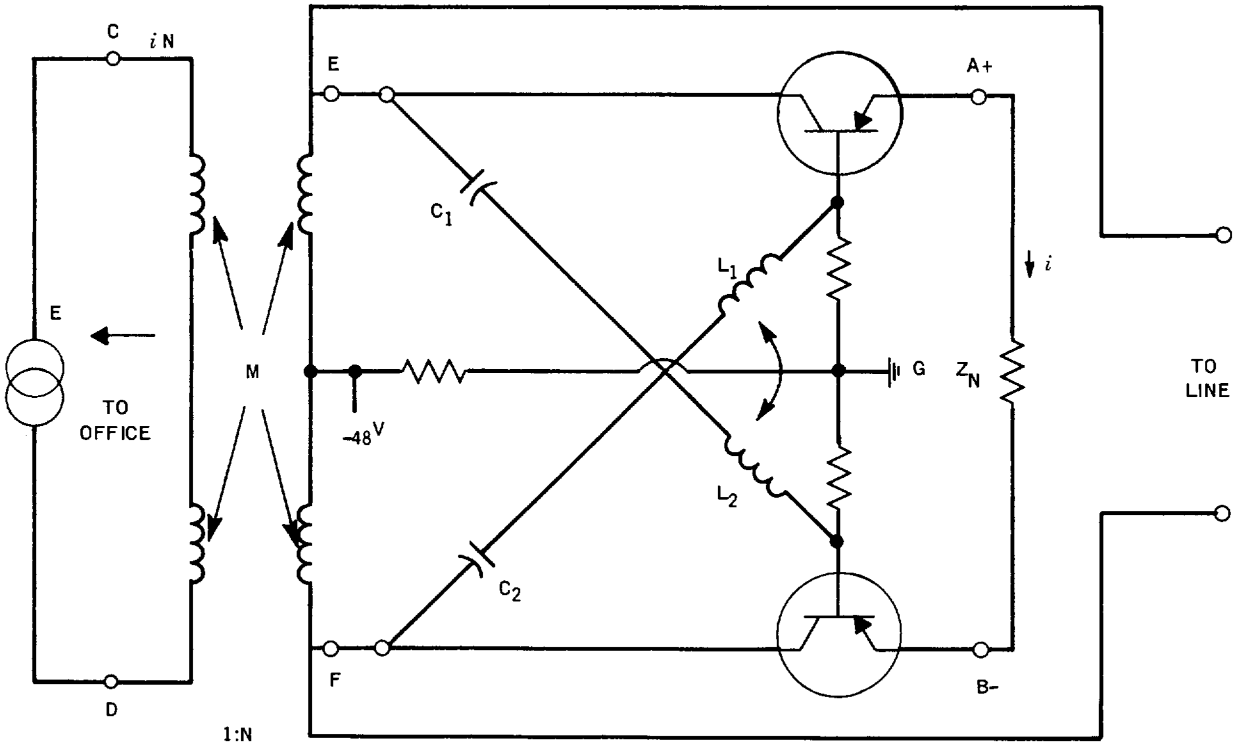
$Y_0 = \frac{1}{Z_0} = G_0 + j\,B_0$ $M = \sqrt{PS}$ $N = \sqrt{\frac{S}{P}}$

Fig. 9



$\frac{Z_N}{2}$
$i$
A
$(K-1)i$
E
$(K-1)i$
$Z_{AB}$
$Ki$
$Ki$
$Z_0$
B
$i$
$i$
LOW INPUT IMPEDANCE
$i$
$\frac{Z_N}{2}$

Fig. 10



C
$iN$
E
TO OFFICE
D
1:N
M
E
F
$C_1$
$C_2$
$L_1$
$L_2$
-48V
G
A+
B-
$i$
$Z_N$
TO LINE

Fig. 11

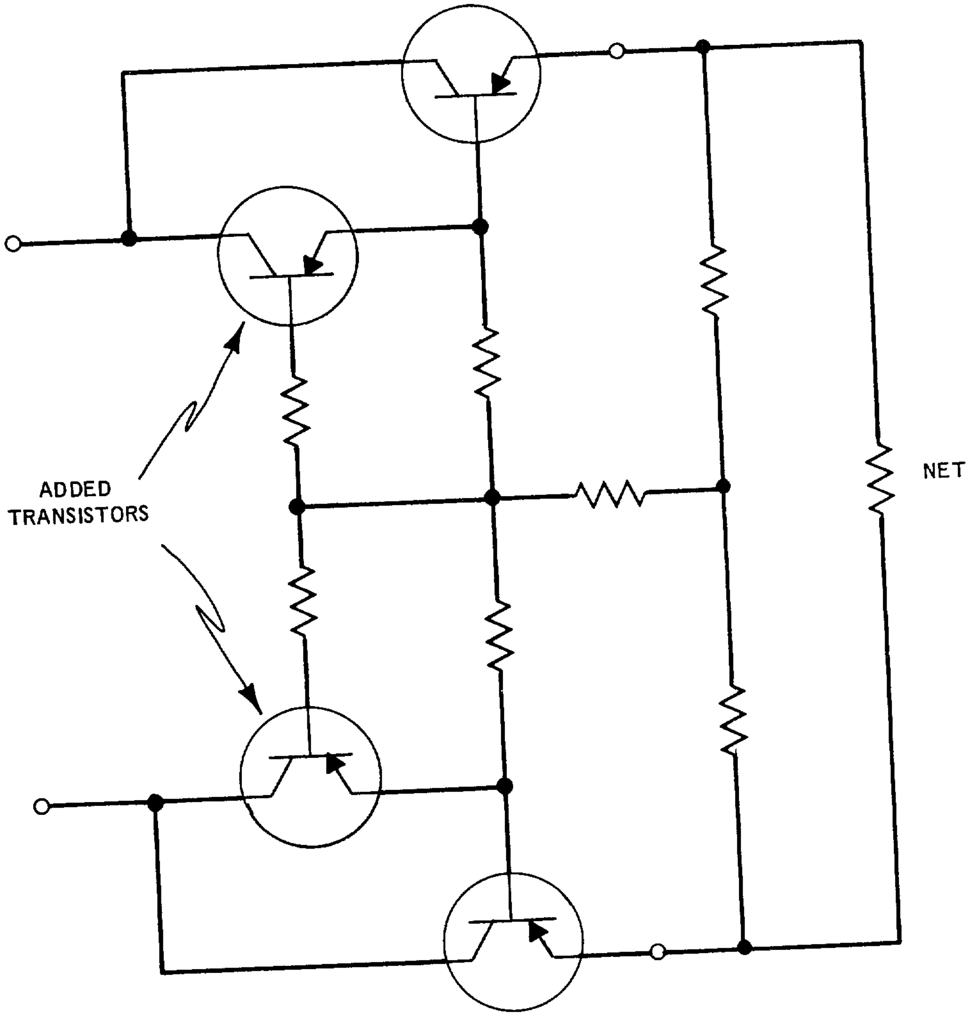


Fig. 12

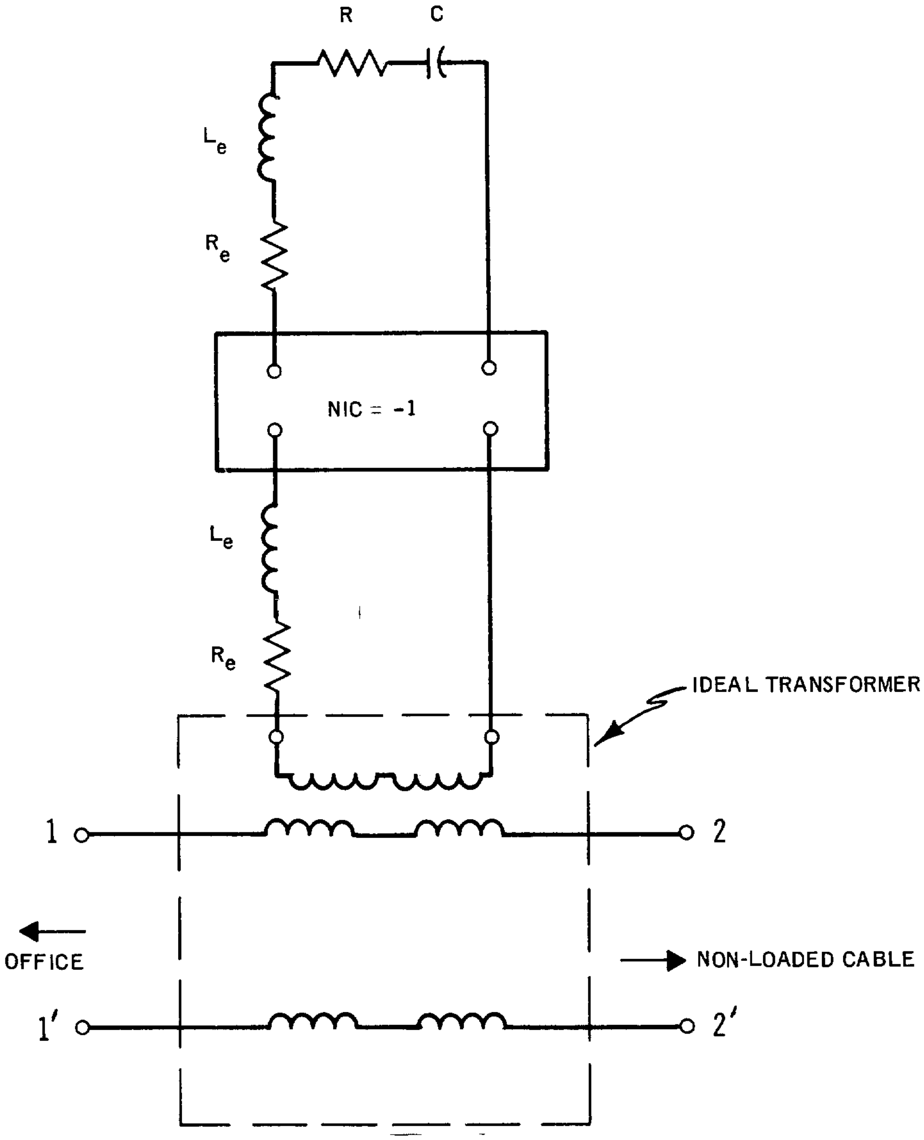

R
C
$L_e$
$R_e$
NIC = -1
$L_e$
$R_e$
IDEAL TRANSFORMER
1
2
OFFICE
NON-LOADED CABLE
1′
2′

Fig. 13