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L5 SYSTEM:

Repeatered Line

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The L5 repeatered line is presented from the viewpoint of a distributed equalization process. Reliable transmission of 10,800 circuits over 4000 miles of coaxial cable with minimum noise is the sole objective of this process. The strategy is to provide equalization in cause-associated increments that place specific bounds on signal-level excursions, thereby insuring an ultralinear, low-noise predictable transmission medium. The individual causes of signal misalignment, both static and dynamic, are examined and the realization of the strategy, which forms a hierarchy of equalizers, is described.

I. INTRODUCTION

The basic function of the L5 repeatered line is to provide a lossless transmission facility between any two L5 terminal stations, which may be located as far as 4000 miles (6400 kilometers) apart, with minimum noise penalty. The objective is to establish and maintain the insertion loss of the 4000-mile line over the message band to be within ±4 dB and the noise to be less than 39.4 dBrnc0 in any of the 10,800 channels transmitted.

The transmission medium for the L5 line is the standard Bell System 0.375-inch disc-insulated coaxial cable.[1] This cable has a loss* that can be conveniently expressed as follows:

$$\text{Cable loss (dB/mile)} = \left[A\left(1 + \frac{0.0062}{\sqrt{f}}\right)\sqrt{f} + Cf\right] + [(T - T_o)D\sqrt{f}], \quad (1)$$

* A 2.1-percent factor has been included in this loss to account for miscellaneous factors such as stranding.

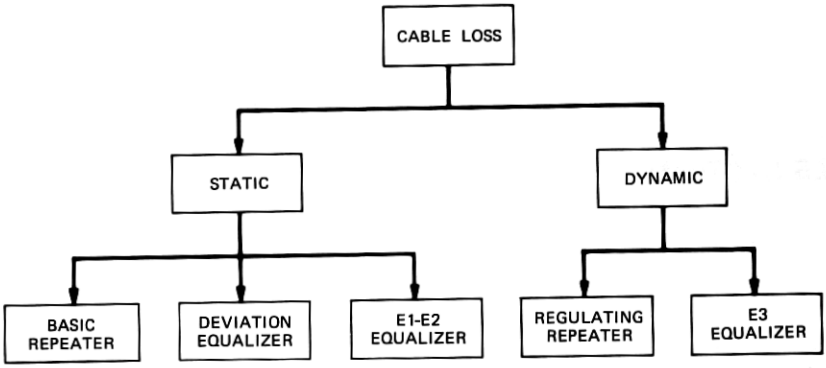

Fig. 1—Equalization strategy.

where the average values of the parameters are

$$\begin{aligned} A &= 3.9002 \text{ dB/(mile MHz}^{\frac{1}{2}}) \\ C &= 0.0047 \text{ dB/(mile MHz)} \\ D &= 0.0043 \text{ dB/(mile MHz}^{\frac{1}{2}}\ {}^\circ\text{F)} \\ T_o &= 55^\circ\text{F}, \end{aligned}$$

and f is frequency in MHz and T is cable temperature in °F.

There are two frequency bands of interest. The first is the message band from 3.1 to 60.6 MHz over which the 4000-mile objectives must be met. To meet these long-haul objectives, however, and to accommodate other system functions, a wider band from 1.6 to 66 MHz is controlled. In eq. (1), the loss associated with the first bracketed term is a function of frequency only. This is called the static loss. The second bracketed term is not only a function of frequency, but also of cable temperature and, hence, of time. This component of cable loss is called the dynamic cable loss.

The strategy to equalize both static and dynamic losses is to provide a cause-associated equalization hierarchy—that is, the various equipment making up the L5 repeatered line each compensate for a particular cause of line misalignment. This relationship is shown in Fig. 1.

This paper discusses the misalignment causes and the associated equipment design and realization.

1.1 Static equalization

Static equalization of the cable is provided in three levels. The first is provided by the basic repeater. With the repeaters spaced at 1-mile intervals, the gain required to match the static component of the cable attenuation given in (1) varies from 4.97 dB at 1.6 MHz to 32.04 dB at 66 MHz.

The second level is provided by the deviation equalizer. It is neither economically nor technically feasible to design the basic repeater with a gain response that exactly matches the cable loss function. The realization of the repeater design results in a slight mismatch between the average basic repeater gain and the average cable loss. This mismatch is referred to as the average design error. An objective was established to hold this error to less than ± 0.1 dB over the frequency band in a 1-mile section or to less than ± 7.5 dB in a 75-mile power-feed section. Compensation for this misalignment is provided by the deviation equalizer located at the transmitting and receiving ends of a power-feed section and in midspan.

Even if it were possible to exactly match the nominal attenuation of the cable by the gain of the basic repeater, there would still be appreciable misalignment in line sections involving many repeaters. This is due to the statistical manufacturing deviations in repeaters and cables, and is impossible to predict for a given line section. The resulting misalignment from both causes is referred to as static line misalignment or static deviation. Equalization of this static deviation forms the third level of static equalization in the L5 system, and is accomplished with adjustable equalizers referred to as E1 and E2 equalizers.

Specific causes for the statistical manufacturing deviations are known. Repeater gain deviations are the result of component and assembly tolerances, while cable loss deviations are the result of variations in the copper conductivity, cable geometry, and dielectric disc conductivity. Copper conductivity and cable geometry influence the A parameter of the cable loss equation and typical variations (one sigma) resulting from this effect are 6 dB at 66 MHz over 75 miles. Dielectric disc conductivity, on the other hand, affects the B parameter of the cable loss equation and amounts to a 1.5-dB change for a one-sigma variation at 66 MHz over 75 miles. However, changes in disc material have occurred since the initial cable production some 25 years ago. Measurements made on a number of these earlier cables indicate a variation in cable loss of about 23 dB at 66 MHz per 75 miles. Since, in addition to new installations, L5 is also intended as a retrofit system for some of these earlier cable applications, this variation becomes important.

1.2 Dynamic equalization

Equation (1) shows that the cable attenuation varies as the temperature of the cable changes. In the L5 system, the cable is buried four feet beneath the ground surface, where the maximum temperature deviation expected in the United States is $\pm 20°F$. This results in a 52-dB change in cable attenuation at 66 MHz for 75 miles. Such change in attenuation is compensated for by repeaters that are called regulating repeaters. These repeaters automatically compensate for changes in the cable attenuation by sensing the temperature of the earth at cable depth and by detecting the level of a pilot tone which is inserted into the message signal at L5 main stations.

This is not the only component of dynamic misalignment that must be equalized, however. Since it is not possible to match exactly the dynamic change of the cable over the entire frequency spectrum, there will be a slight misalignment or slight difference between the change in gain of the regulating repeater and the change in loss of the cable, referred to as the tracking error. In addition, slight differences occur between the change in gain of the basic and regulating repeaters themselves as the temperature of their environment changes. These two effects, taken together, are referred to as the residual dynamic transmission deviation. While over any short-line section this deviation is very small, the accumulated effects over a 75-mile section are typically ± 3 dB at 66 MHz. This effect is automatically equalized by an E3 equalizer. The time-varying transmission deviations of an L5 line are detected by four pilot tones spaced across the line frequency spectrum. Four networks in the E3 equalizer automatically respond to the pilot levels to correct the residual dynamic transmission deviation.

The remaining sections of this paper describe in more detail each level of equalization in the L5 system.

II. BASIC REPEATER

2.1 Basic repeater function

The function of the basic repeater is to provide a fixed gain to compensate for the attenuation of 1 mile of coaxial cable. Its characteristics have the dominant effect on the overall system performance. Thus, the requirements it must meet are stringent to assure that the overall 4000-mile system objectives are met. These requirements cover numerous areas such as specification of gain match to cable loss, intermodulation distortion, power-handling capacity, noise figure, return loss, and temperature coefficient. The simultaneous achievement of all these specifications is a major accomplishment in the design of the L5 system.

2.2 Basic repeater design requirements

Two of the most important design requirements are (*i*) that the average gain of the basic repeater must match the loss (at 55°F) of one mile of coaxial cable over the message band to within ±0.10 dB and (*ii*) that the two-sigma limit on the distribution of repeater gains must not differ from the average characteristic by more than 0.10 dB. These requirements necessitate an amplifier design with an overall square-root-of-frequency gain shape of 6.91 dB at 3.1 MHz to 30.69 dB at 60.6 MHz.

The second- and third-order interchannel modulation distortion (hereafter referred to as intermodulation noise) of the repeater must be extremely small. The second-order intermodulation coefficient objective varies from −105 dB at the low end of the spectrum to −70 dB at the high end. The third-order intermodulation coefficient objective varies from −128 dB at the low end to −110 dB at the high end. These objectives are derived from calculations involving the repeater noise figure, output spectral density of the signal, and the assumed law of addition, from repeater to repeater, for intermodulation noise.

To avoid a rise in system noise because of peak busy-hour message load, the basic repeater must be capable of sustaining a load of +24 dBm before overloading. For the L5 system, the conservative definition of overload is used as the point where the modulation coefficient degrades by 0.5 dB. An ultralinear high-power amplifier is obviously required as the result of the overload and intermodulation objectives.

Noise figure of the repeater obviously affects system signal-to-noise performance and therefore should be kept to a minimum. The noise figure objective is 8.5 dB at the low frequencies and slowly decreases to 5.5 dB at the high frequencies. Contrary to overload and intermodulation, this requirement calls for a low thermal noise amplifier. Additional objectives that influenced the amplifier design were surge protection, return loss, and temperature coefficient.

To summarize, the basic repeater objectives dictate a shaped-gain, low-noise, high-power ultralinear repeater. One method of achieving the shaped-gain requirement would be to design a flat-gain amplifier and a separate network for the shaped loss. The shaped loss would be equal to the maximum gain required minus the minimum gain required (for signal transmission). This amount of loss (25.13 dB), if placed at the input, would lead to an excessively high noise figure at low frequencies; if placed at the output of the amplifier, this loss would excessively penalize intermodulation distortion and overload. On the other hand, these penalties may be avoided by designing the amplifier with a square-root-of-frequency shape ($\sqrt{f}$) network in the feedback path, thus creating a shaped-gain amplifier. As a result, additional negative feedback is available; it improves the intermodulation distortion and gain deviations resulting from μ-path variations and improves return loss. However, the shaped-gain feedback amplifier must be properly compensated to avoid high-frequency stability problems.

2.3 Basic repeater realization

The realization of the basic repeater includes two feedback amplifiers and a number of passive networks shown in block diagram form in Fig. 2. Both amplifiers have shaped loss networks in their feedback paths to accomplish the $\sqrt{f}$ gain required. Details of the amplifiers and other networks are given in the sections that follow.

2.3.1 Twin jack and bridging pad

The twin jacks, shown at the repeater input and output, are located behind the mounting surface for the repeater in the manhole. Their purpose is to provide connection from the coaxial line to the repeater with an additional port exposed for outside connection. The outside connection serves two purposes. First, it allows injection of fault-locating tones through bridging pads to the repeater input and output. Second, it allows for "power patching" of repeaters. This is done by removing the bridging pads and replacing them with a coaxial cable patch. The repeater may then be removed and replaced without interrupting the coaxial line dc power. This simplifies and expedites repeater replacement.

The bridging pad has an impedance of about 2000 ohms facing the twin jack. Since it connects with a coaxial plug, there is also a parasitic capacitance to ground. To maintain the 75-ohm impedance required at the repeater input and output, an inductor was designed into the twin jack. This is shown schematically in Fig. 3. By choosing L such that $\sqrt{L/C} = 75$ ohms, a lumped section of 75-ohm coaxial cable was approximated, and the impedances were maintained. The shunting effect of the 2000-ohm resistor is compensated for by the impedance of the low-frequency networks that are discussed in Section 2.33.

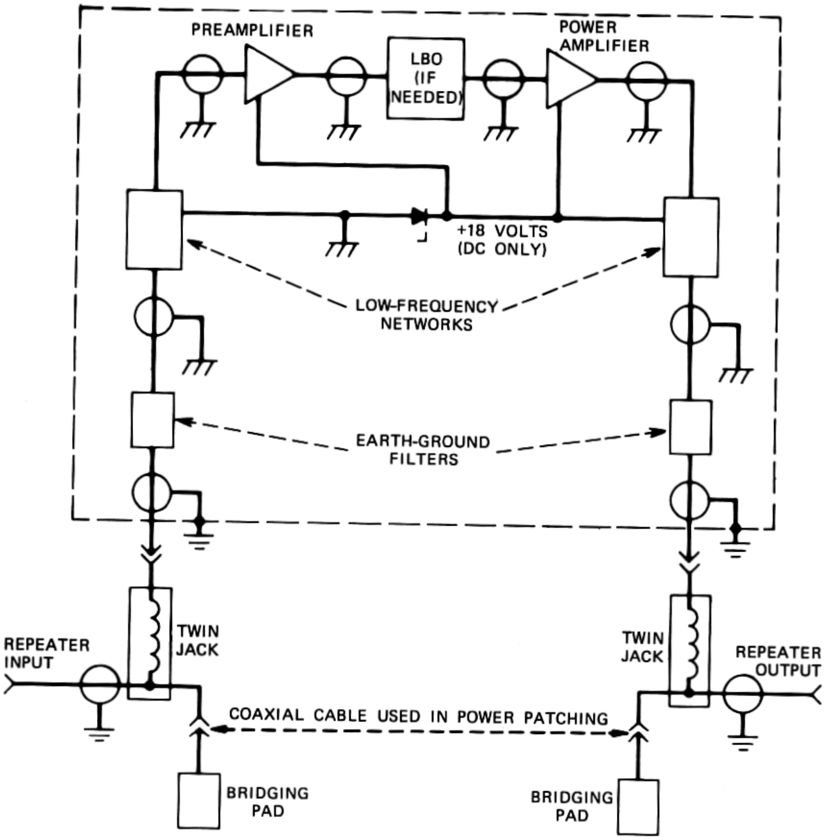

Fig. 2—Basic repeater.

The effect of the twin jack and bridging pad are included in the performance specifications of the repeater.

2.3.2 Earth-ground filters

The earth-ground filters, shown at the repeater input and output, consist of sections of high-voltage capacitors and ferrite beads which form a filter. They provide transmission of signal and power through the outer wall of the repeater (which must be kept at earth ground for safety reasons) to an inner "floating ground." Signals developed across the high-voltage output capacitors are then sufficiently attenuated by the filter so as not to couple into the input high-voltage capacitor, and vice versa. Failure to provide this filtering action would result

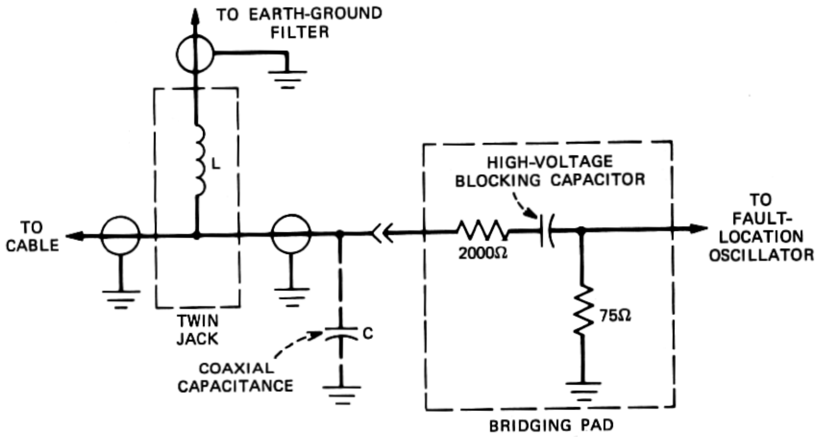

Fig. 3—Twin jack.

in signals coupling from output to input, which produce unequalizable ripples across the frequency spectrum of the repeater. The earth-ground filters are the only high-voltage components in the repeater; all other components operate with respect to the internal "floating ground."

2.3.3 *Low-frequency networks*

The two low-frequency networks (see Fig. 2) differ slightly in internal component choice, but perform essentially the same functions. The original purpose of the low-frequency networks was to provide some loss at the very low frequencies of the L5 spectrum to compensate for excess gain of the amplifiers. However, during the evolution of the repeater design, other functions were added to the low-frequency networks. A simplified schematic of a low-frequency network is shown in Fig. 4. The network, comprised of Z1, Z2, and Z0, provides the low-frequency loss shaping previously mentioned. Resistor R1 provides both an impedance match to 75 ohms (in conjunction with the 2000 ohms in the bridging pad) and a current limiting function for surge protection. It has a small flat loss effect on the repeater response, which is compensated for by the main amplifiers.

G1 is a gas tube surge protector having a dc voltage breakdown of about 90 volts and, under normal circuit conditions, introduces only 1.5 pF of shunt capacitance. Inductor L1 provides for the separation of the dc line current from the L5 message signal. It maintains a high reactance over the frequency band to keep signal losses to a minimum.

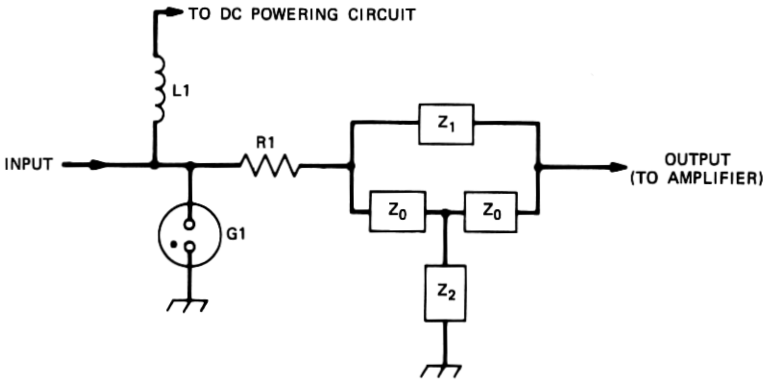

Fig. 4—Low-frequency network.

2.3.4 Line-build-out network

The line-build-out network (LBO) is a passive network that provides loss equivalent to a given length of coaxial cable. It is used to reduce the repeater gain from the equivalent of 1.0 mile of cable down to 0.5 mile in steps of 0.1 mile. This allows flexibility for those cases in which the sections of cable are shorter than 1.0 mile (usually for easy access to the repeater location or because of a physical obstruction in the location of the manhole). Since the line-build-out network is an integral part of the basic repeater, a separate repeater identification (code) must be associated with each of the six line-build-out networks. This is in contrast to earlier L-carrier systems in which the line-build-out network was provided to the field as a separate network. The advantage of this approach is that it allows for precise control of the gain interaction between the line-build-out network and the input impedance of the power amplifier.

2.3.5 Amplifier design considerations

To achieve the required gain, noise figure, and distortion requirements of the basic repeater, new ultralinear wideband transistors were required. The characteristics of these transistors and other transistor design considerations are covered in a companion article.[2]

As previously mentioned, each amplifier uses negative feedback. Good stability of the amplifier requires minimization of the length of the $\mu\beta$ path, which was achieved by employing thin-film technology with appliqued components, yielding a tri-level circuit realization (see Fig. 5). The three layers thus formed are comprised of (*i*) thin-film

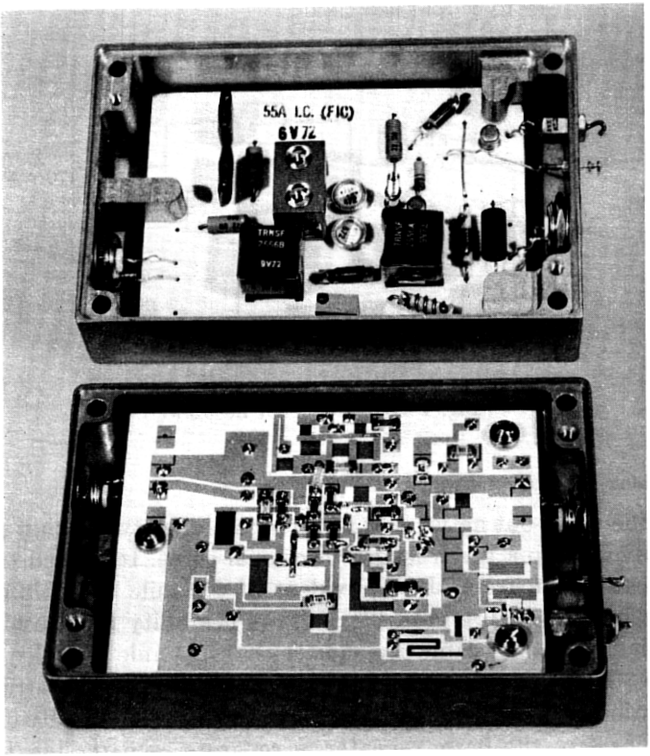

Fig. 5—Tri-level preamplifier.

resistors and metal system interconnections, (*ii*) discrete leaded components (mostly inductors, transformers, and transistors) and (*iii*) leadless ceramic chip components (capacitors and thick-film resistors). This tri-level circuit minimizes the $\mu\beta$ path length and increases the available feedback. In addition, since thin-film resistors and ceramic chip capacitors have no external leads, parasitic effects are minimized, thus improving the circuit performance at the high frequencies required in the L5 system.

The feedback networks of both amplifiers required precision components to achieve the required gain characteristic. An initial tolerance of ± 0.1 percent is maintained with the thin-film-shaping resistors, and selection limits of ± 0.03 dB flatness across the L5 spectrum are required on the transformers.

Other details of the physical realization of the repeater are discussed in more detail in a companion article.[3]

2.3.5.1 Preamplifier. The preamplifier is a negative feedback, hybrid-input, hybrid-output amplifier depicted in Fig. 6. This circuit configuration was chosen, after careful computer simulation of various alternatives, for its low-noise figure while still maintaining good intermodulation performance and stability. The input and output hybrid transformers have impedance ratios of 75:65 + 28 ohms. This ratio was chosen to minimize the amplifier noise figure, while still maintaining physical realizability and reproducibility of the hybrid. Decreasing the "through loss" of the hybrid to decrease the noise figure results in increasing the loss through the $\mu\beta$ ports of the hybrids, which then results in an increased minimum gain of the amplifier. It is this minimum gain, resulting from the hybrid effect, that is compensated for by the low-frequency networks.

The input stages were chosen as common emitter, biased at 30 mA and 5 volts, to achieve a low noise figure and to provide sufficient gain to minimize the effect of the noise figure of the output stages on the overall amplifier noise figure. The output stages are biased at 110 mA and 12 volts for optimum intermodulation performance. Both input and output stages consist of two transistors connected in parallel to reduce the intermodulation noise as follows. The input signal current divides equally, and each transistor then carries one-half the total current. The controlling third-order nonlinearity is primarily current-dependent and, as a result, the output distortion of each transistor is reduced to one-eighth $[(\frac{1}{2})^3]$. When combined from both transistors, the output distortion is one-fourth that of a single stage, or a 12-dB improvement. Similarly, for second-order distortion, a 6-dB improvement is achieved. To achieve these improvements, the transistors are paired by matching their current gain.

Two adjustable elements are in the preamplifier—a capacitor and a potentiometer. Each has a very limited range and is used for factory adjustment of the gain response, particularly at high frequencies. Computer sensitivity runs were used to choose the variable elements for matching the normal amplifier gain shape variation. The capacitor affects the balance between loop and local feedback and hence influences the closed loop gain through the $\mu\beta$ effect,[4] which varies the repeater gain only at the very high frequencies. The potentiometer is in the bridged-T β-network. Changes in its resistance directly vary the β loss and, hence, the amplifier gain. This adjustment changes a broad, high-frequency shape and is made after the total repeater has been assembled, to set the gain at 42.880 MHz (the temperature pilot) to within ±0.01 dB of the nominal value. This precise gain adjustment

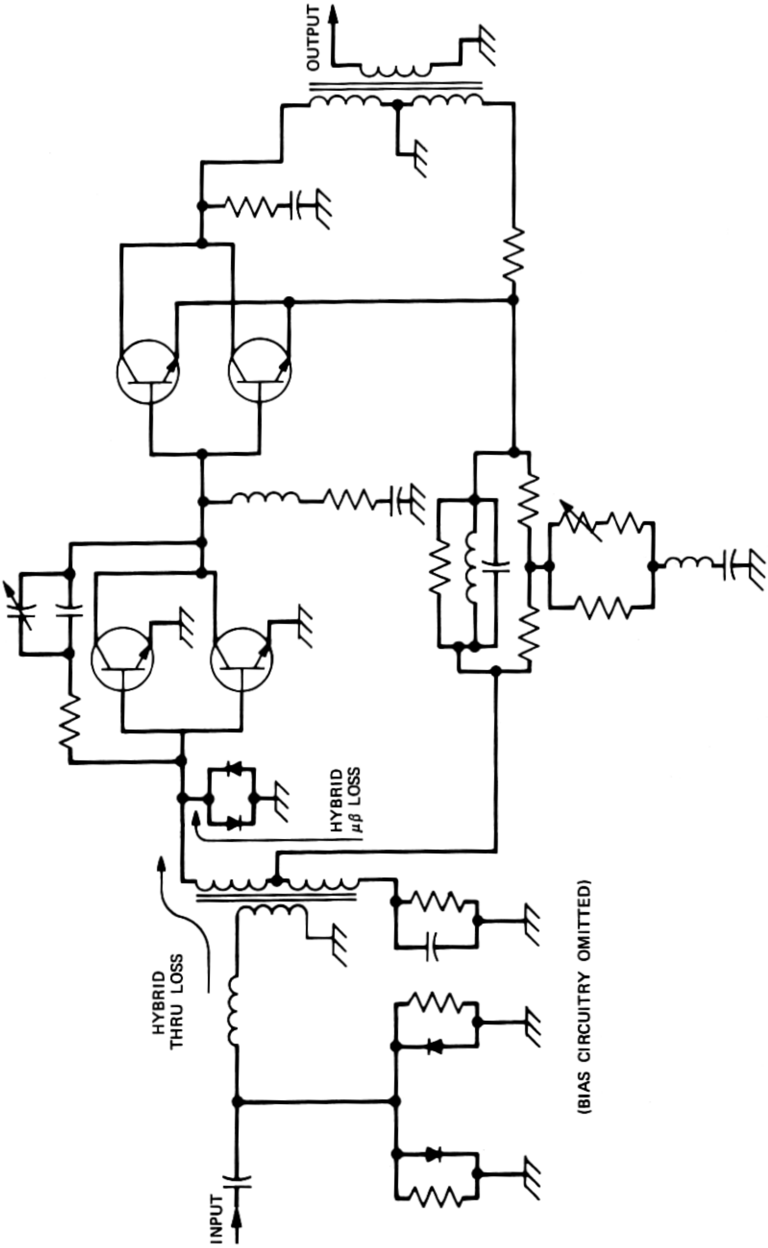

Fig. 6—Preamplifier schematic.

minimizes gain error introduced by the basic repeater at the temperature pilot frequency and therefore reduces the range requirements of the regulating repeater.

2.3.5.2 Power amplifier. The most difficult objective to meet in the basic repeater was the intermodulation distortion. After an extensive program of computer modeling of transistor and circuit nonlinearities[5] and laboratory evaluation, the circuit topology shown in Fig. 7 was chosen for the power amplifier. It has shunt feedback at the input and hybrid feedback at the output. The μ path is a common emitter-transformer-common-base arrangement. The interstage transformer provides current gain to minimize the effect of distortion in the common emitter stage and also provides a more nearly optimum set of interface impedances between the stages. An autotransformer was chosen as the output hybrid to obtain maximum bandwidth and minimum phase shift.

As in the case of the preamplifier, all transistors are paralleled for improvement in distortion. The bias point of 15 volts and 110 mA was chosen to minimize the third-order distortion.

Computer simulation revealed that the closed-loop gain at high frequency was sensitive to the capacitance at the common base output node. This node capacitance varies as a result of variation in transistor capacitance, surge diode capacitance, choke winding capacitance, and circuit parasitic capacitance. Because these capacitances cannot be controlled to the desired tolerance, an adjustable capacitor was added.

To meet the distortion requirements, the loop feedback was maximized. To maintain stability margins at the maximized feedback, an input hybrid was not used. The input network serves to form part of the overall amplifier gain shaping, but at the expense of presenting a nonconstant impedance to the preamplifier (or LBO, if used). This impedance mismatch creates a gain interaction when line-build-out networks are placed between the preamplifier and the power amplifier. To offset this interaction, power amplifiers intended for use in repeaters with line-build-out networks are adjusted to a different nominal gain.

2.3.6 Surge protection

Buried cable systems, although shielded, are still subjected to a number of transients induced from within and without the system. These include (*i*) high-voltage line turnup,[6] (*ii*) power patching and subsequent insertion of an uncharged repeater into the high-voltage line (*iii*) lightning, and (*iv*) 60-Hz induction. To protect the repeater

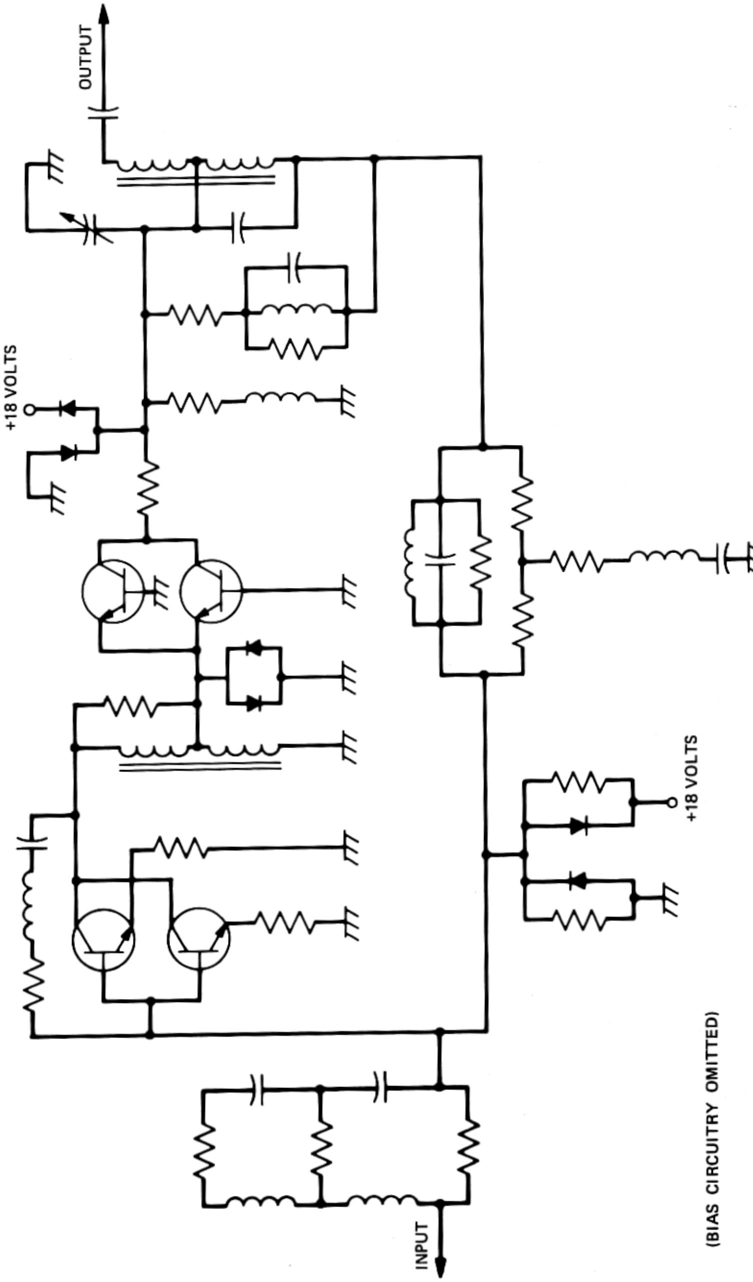

Fig. 7—Power amplifier schematic.

from these effects, both primary and secondary protection are employed. Primary protection consisting of a gas tube protector and series resistor was described in Section 2.3.3. Secondary protection is included in both the preamplifier and the power amplifier in the form of surge-rated input- and output-coupling capacitors as well as surge-rated diode pairs. Similar surge-protecting circuitry is included in the regulating and equalizing repeaters.

2.3.7 Basic repeater performance

One performance feature that must be determined very early in the development of a coaxial system is the match of the gain of the line repeaters to the loss of the cable. This information is necessary for the design of the deviation equalizer, which is described in the next section. Although prototype information was available on the degree of match, the results of actual production runs with final component types are required to establish a sufficiently accurate frequency response of the repeaters. These data were obtained with the cooperation of the manufacturer, Western Electric, from measurements of the early production repeaters by Bell Laboratories personnel. The details of the accuracy and environment of these measurements are described in a companion paper.[5] The results of these measurements, when compared to the installed line measurements, established a high degree of confidence in the reproducibility of the design and in the validity of the parametric equation describing the cable loss. The results also provided repeater temperature coefficients which were required for the design of the network shapes in the E3 dynamic equalizer.

The following illustrations demonstrate some performance features of the basic repeater. Figure 8 illustrates the match to cable loss and Fig.

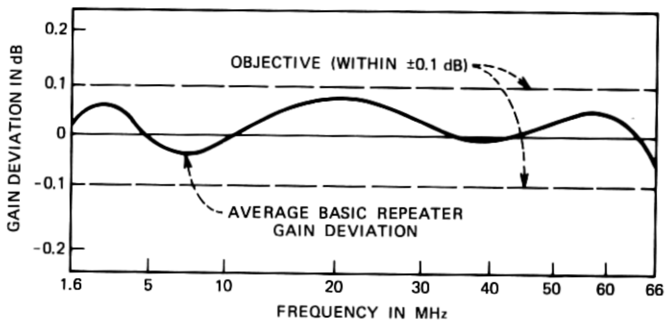

Fig. 8—Average deviation of basic repeater gain from the average loss of 1 mile of coax cable.

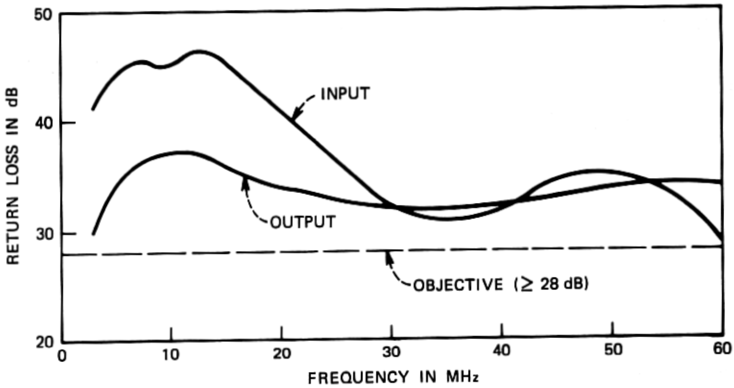

Fig. 9—Basic repeater return loss (average of 2071 production repeaters).

9 represents typical input and output return loss. Overload and temperature coefficients are shown in Figs. 10 and 11, respectively. The noise figure of a basic repeater is plotted in Fig. 12. Second- and third-order modulation are depicted in Fig. 13.

III. DEVIATION EQUALIZER

It has been pointed out that exact equalization of the static attenuation component of the cable loss is not feasible. The design transmission objective of the basic repeater is to match the nominal cable attenuation characteristic to within ±0.1 dB. The difference between

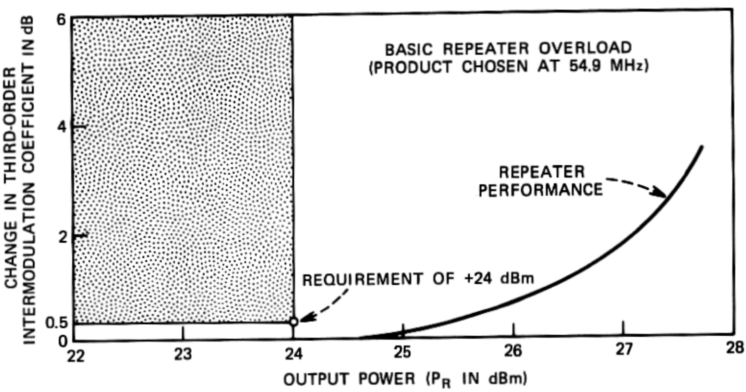

Fig. 10—Basic repeater overload (product chosen at 54.9 MHz).

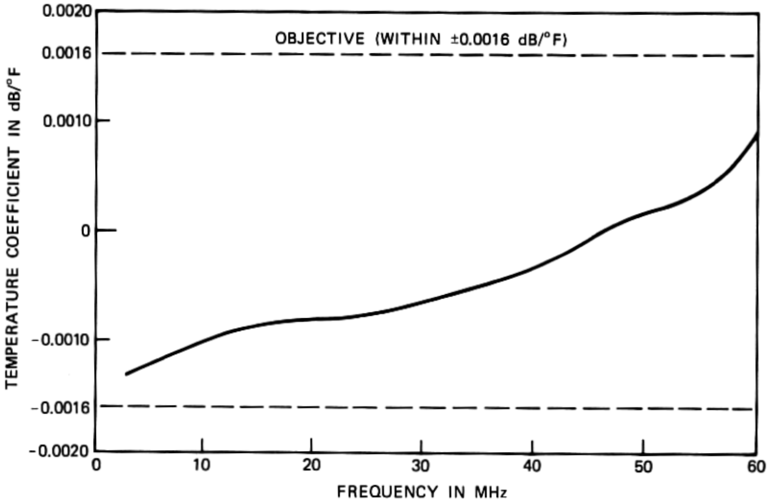

Fig. 11—Basic repeater temperature coefficient.

the gain of the average repeater and the nominal loss of the cable is called the average design error and is obviously predictable. Equalization of the average design error is accomplished by providing fixed equalizers in the transmitting and receiving repeater in the main stations and in the equalizing repeater located at approximately the midspan location between main stations. In previous L-carrier systems, the loss characteristic of the deviation equalizer was determined as

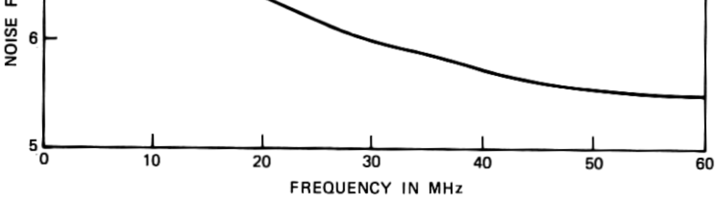

Fig. 12—Basic repeater noise figure.

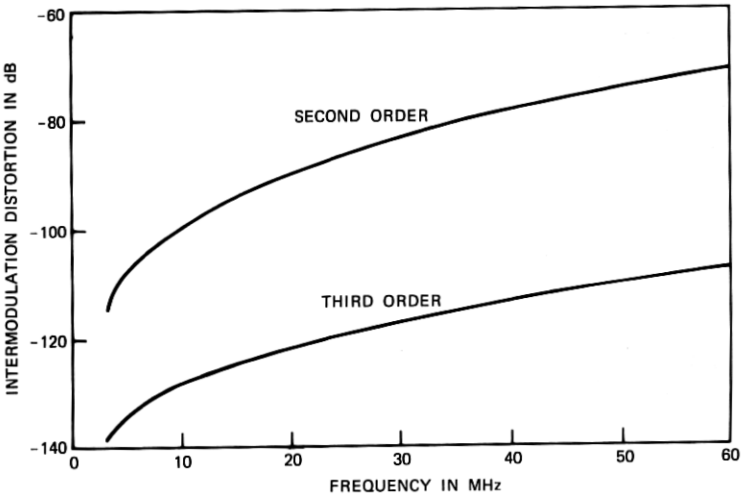

Fig. 13—Basic repeater intermodulation distortion.

follows:

(*i*) Determine the average design error as a function of frequency for each type of line repeater.

(*ii*) Determine the cumulative average design error for the average expected line section length.

(*iii*) Modify the cumulative design error to compensate for any design error at the temperature pilot frequency by adding (or subtracting) the function $K\sqrt{f/f_T}$, where K is the cumulative design error at the temperature pilot frequency, f_T. This is necessary because any gain offset at the pilot frequency will be operated on by the regulating repeaters as if it were caused by a line temperature change.

(*iv*) Apportion the resultant characteristic, which is the desired total insertion loss characteristic (inverted in sign, of course) of all line deviation equalizers, equally to all equalizers.

(*v*) Realize a network to match the desired characteristic according to some error criterion (usually a minimum mean-squared criterion).

In any actual line section, a residual gain or loss characteristic will still remain, since the line will not contain all average repeaters and cable, and will be of a different than average length. It is difficult to determine beforehand whether this residual characteristic is equalizable within the capabilities of the variable equalizers provided in the system.

In the L5 system, a different design approach was taken. Rather than the deviation equalizer being designed to match the average design error, it was designed by a computer program so that the difference between the deviation equalizer and the average design error is optimally equalizable by the L5 adjustable equalizers.

The results can be seen in Fig. 14. Curve (a) shows the average design error of the average L5 power-feed section consisting of 60 basic and 10 regulating repeaters. Curve (b) shows the difference between the average design error and the deviation equalizers. Curve (c) shows the equalized characteristic predicted by the computer program using mathematical descriptions of the equalizers.

This approach has another advantage in terms of the system development time. Repeater development and variable equalizer development both require long intervals. Using the former approach, optimum variable equalizers cannot be developed concurrently with the repeaters. In the new approach, the deviation equalizer is used to optimally "marry" the repeater design error and the equalizers. The development time of the fixed deviation equalizer is very short compared to the repeaters and variable equalizers.

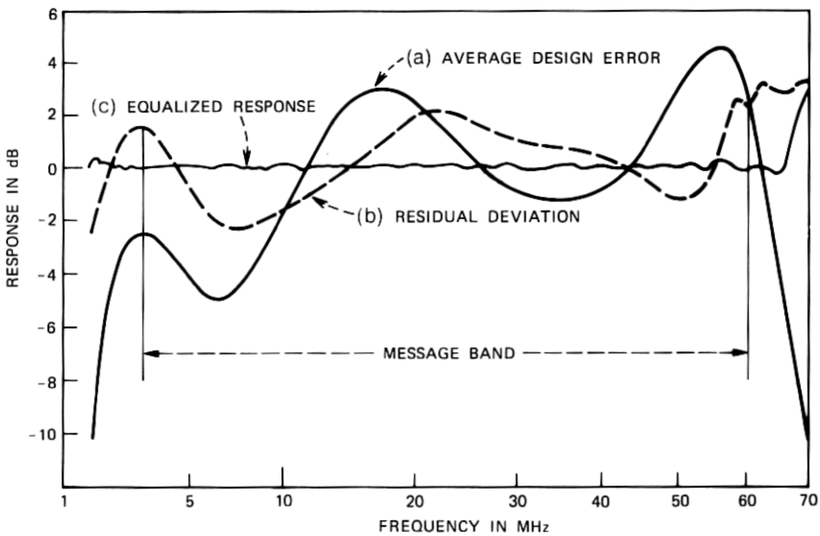

Fig. 14—Static equalization.

IV. ADJUSTABLE STATIC EQUALIZERS

As indicated in the previous section, gain deviations of the line repeaters resulting from manufacturing differences from repeater to repeater and differences in cable loss from section to section, taken together, result in a static misalignment of a line section. Since there are a large number of contributors to this static misalignment, misalignment of any line section as a function of frequency is extremely difficult, if not impossible, to predict *a priori*. An equalizer designed to compensate for this misalignment must therefore be able to accommodate a wide range of shapes. This implies an adjustable equalizer with considerable flexibility.

On the other hand, the static misalignment of any line section will be a very slowly varying function of time. These considerations led to the design of manually adjusted equalizers that are adjusted upon initial installation and, at very infrequent intervals thereafter, governed by such factors as route growth and line equipment replacement.

In-service versus out-of-service adjustment is another important consideration affecting adjustable equalizer design. While in-service equalization is preferable from an operational standpoint, the concomitant constraints imposed on the location of equalizer adjustment tones in the frequency spectrum make the achievement of an optimum equalizer setting difficult. Since the equalizers will be adjusted infrequently, however, out-of-service adjustment imposed no system penalty and, in fact, is considered necessary to achieve the objective of ± 0.4 dB in a switching section.

In wideband coaxial systems, two types of equalizers are commonly used for amplitude equalization—transversal equalizers[7] and Bode

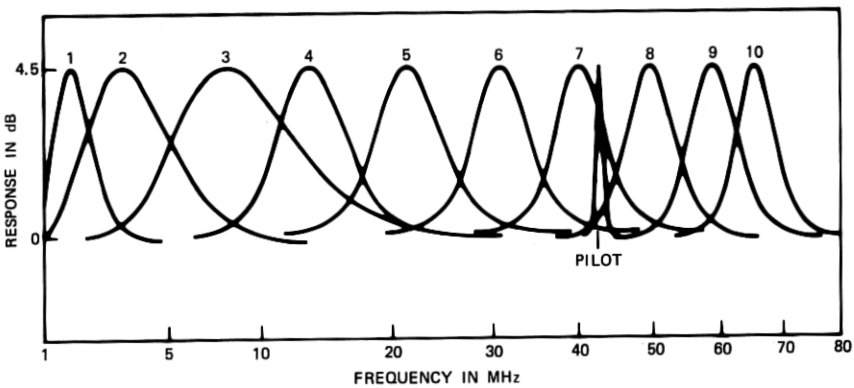

Fig. 15—E1 equalizer shapes.

equalizers.[8] After extensive studies conducted early in the L5 development, Bode equalizers were selected primarily because the algorithm for adjusting the equalizer setting to match the shape of the line is simpler and converges more rapidly to the optimum settings than that for transversal equalizers. This is an important consideration. Achieving the 4000-mile equalization objective depends to a large degree on the ability of the telephone company craftsperson to achieve the optimum equalizer settings within the individual line sections. Therefore, it is a distinct advantage to have a simple algorithm. Furthermore, a simple algorithm is much more amenable to mechanization. Mechanization, in turn, insures the achievement of optimum equalization. This is discussed in detail later.

In the L5 system, two manually adjustable equalizers, the E1 and E2 equalizers, are provided to compensate for the residual static misalignment of the L5 line sections.

4.1 E1 equalizer

The E1 equalizer contains ten broad shapes spaced across the L5 frequency band from 1.6 to 66 MHz plus one very narrow shape centered at the L5 temperature pilot location. They are shown in Fig. 15. In each L5 section are three E1 equalizers, one in the transmitting main station, one in the receiving main station, and a third at the equalizer manhole. From a functional standpoint, the three E1 equalizers are identical. The manhole E1 equalizer differs from the others in that it is packaged to fit in a manhole apparatus case and also in that it derives its power from the dc current on the center conductor of the coaxial cable. The primary function of the E1 equalizer is to provide relatively coarse equalization of the L5 line and to contain the message signal within relatively narrow levels as it traverses the line so that excessive noise penalties will not accrue.

The nominal insertion gain of the E1 equalizer—that is, the gain of the equalizer when all bumps are in their reference state—is 0 dB. The gain or loss around zero of each bump can be set independently of any other bump by means of a manual control on the front panel of the equalizer. Note that, with the exception of the pilot bump, each bump shape has a zero crossover close to the center frequency of the bump immediately preceding it. Taken collectively, we refer to this property of the E1 bumps as being one-way orthogonal with respect to their center frequency. The significance of this property is that, if the required gain setting of bump i is determined by a measurement of the line with only bumps $i - j$, $j = 1, 2, \cdots, i - 1$

previously set, then the required gain will not be altered by the subsequent setting of bumps $i + j$, $j = 1, 2, \cdots, N - i$ (where N is the total number of bumps in the equalizer). It is this property that allows rapid convergence of the equalizing algorithm.

The pilot bump has a very specific purpose. The transmission level layout of the L5 line is such that the absolute power level of the temperature pilot at the input to the E1 equalizer is always the correct level. If the E1 equalizer were to change the pilot level (by insertion of gain or loss by bump number 7, shown in Fig. 15, for instance), then the regulating repeaters in the line following the E1 equalizer would act to correct the change with $\sqrt{f}$ shape. In general, this would prevent the equalization from converging and the line could not be equalized. The purpose of the pilot bump, then, is to maintain the pilot leaving the E1 equalizer at its correct transmission level regardless of any other gain setting within the equalizer. The pilot equalizer is narrow enough so that it does not affect the equalization of the message band surrounding the pilot frequency.

The E1 equalizer is realized by a series of four amplifiers alternately interconnected with four passive networks, as depicted in Fig. 16. Two networks are series double-bump Bode networks and the third is a triple-bump Bode network. The theory of Bode network design and the realization of the networks are discussed in a companion paper.[9] The amplifiers in the E1 equalizer serve several purposes. First, the gain of the amplifier makes up for the flat loss of the networks so that the overall insertion loss of the equalizer is 0 dB. Second, they provide isolation between the Bode networks so that the return loss interaction inherent in these networks is minimized. Finally, each amplifier provides one bump shape by the inclusion of a Bode network in its feedback path. A typical amplifier schematic is shown in Fig. 17.

The most difficult design problem in the E1 (and E2) equalizers was the achievement of the overall frequency response. The gain or loss of each amplifier or network comprising the equalizers is adjusted during its manufacture to be within ±0.05 dB of its respective nominal value. The requirement on the overall equalizer is that, in its reference state, the gain of the equalizer should be within ±0.1 dB of zero when all the Bode networks are in their flat condition. This has been achieved by providing a trimming network adjusted during the assembly of the E1 equalizer to trim the characteristic of the E1 to within the desired requirements. Because the network and amplifier gains are tightly controlled, the trim equalizer is a simple network containing only a flat and a slope term.

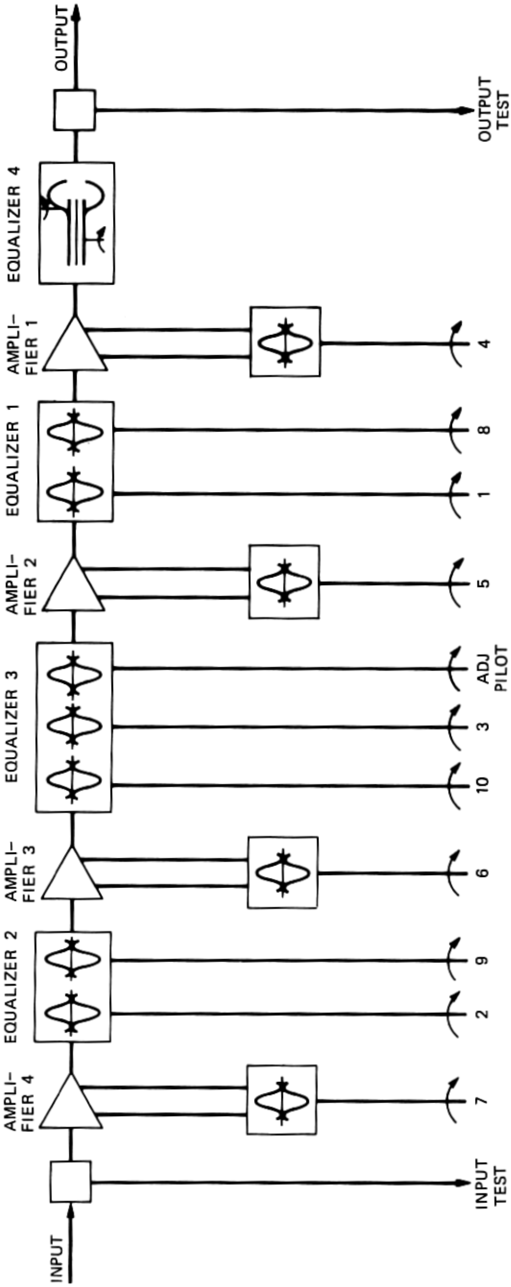

Fig. 16—E1 equalizer transmission circuitry.

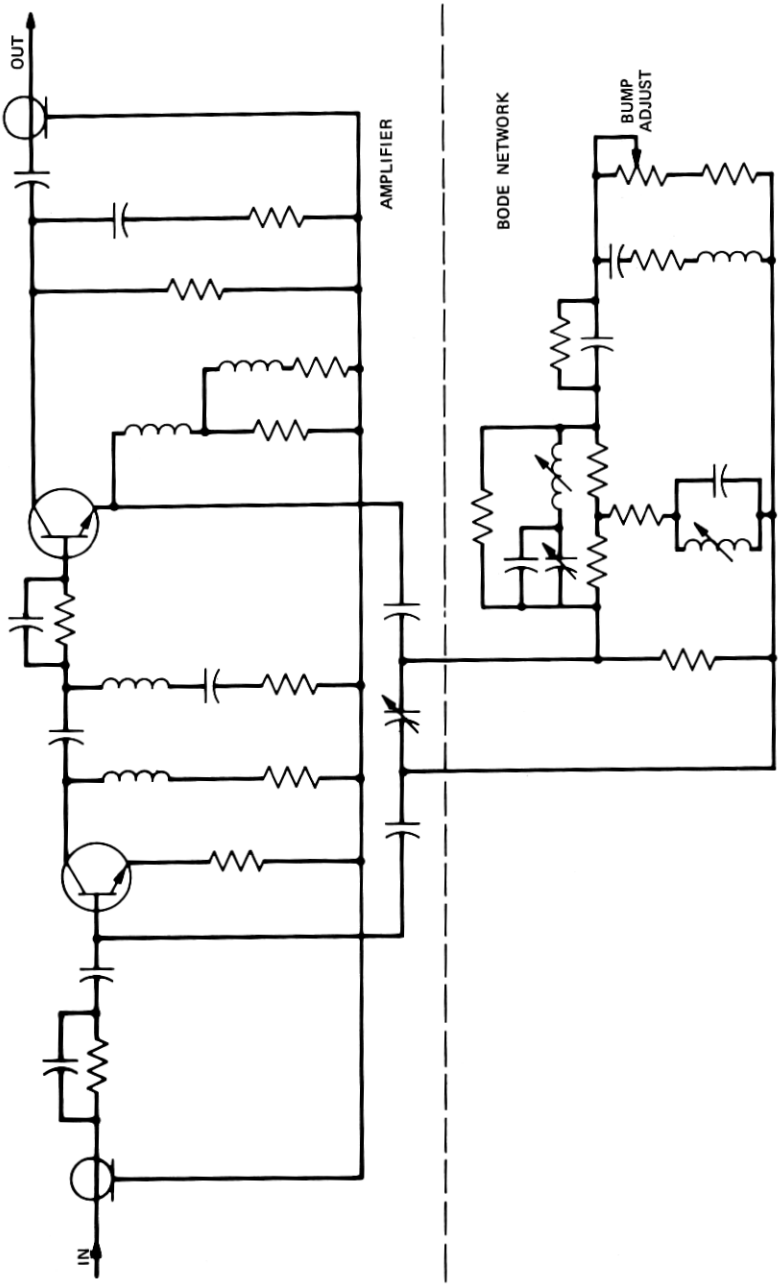

Fig. 17—Equalizing amplifier schematic.

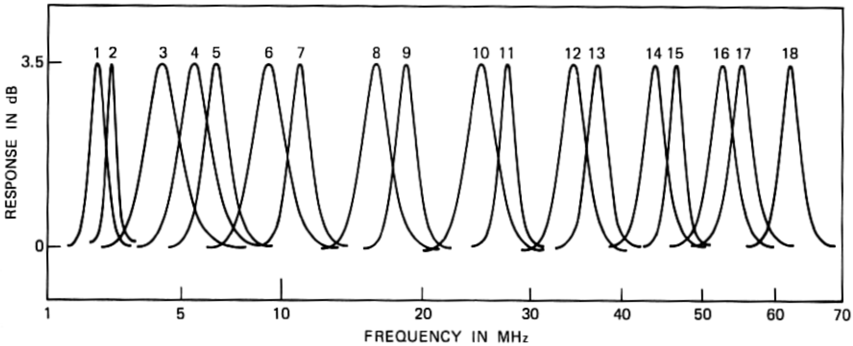

Fig. 18—E2 equalizer shapes.

4.2 E2 equalizer

The E2 equalizer is similar in design to the E1 equalizer. It contains 18 narrow shapes spaced across the 1.6- to 66-MHz spectrum. E2 equalizers are located in main stations at both the transmit and receive ends of a power-feed section. The location of the E2 bumps are indicated in Fig. 18. The primary function of the E2 equalizer is to provide the fine-grained equalization of the L5 signal so that, coupled with the E1 equalizer, the overall equalized response of a switching section is within ±0.4 dB of zero from 1.6 to 66 MHz.

The E2 equalizer contains seven amplifiers alternately interconnected with seven networks. Six networks are series double-bump Bode networks. The seventh network is a trim equalizer similar to the one in the E1 equalizer. Six amplifiers contain Bode networks in their feedback paths, while the seventh is a flat-gain amplifier. Salient features of the E1 and E2 equalizers are shown in Table I.

Table I — Static equalizer characteristics

	E1 Equalizer	E2 Equalizer
Frequency Response (MHz)	1.6 to 66	1.6 to 66
Nominal Gain (dB)	0.0 ± 0.1	0.0 ± 0.1
Input Return Loss (dB)	>30	>30
Output Return Loss (dB)	>30	>30
Number of Shapes	10 Broadband 1 Pilot Adjustment	18
Range of Shapes (dB)	±4.5 Broadband ±1.0 Pilot Adjustment	±3.5
Number of Amplifiers	4	7
Number of Networks	4	7
Full Load Output Power (Single Sinusoid) (dBm)	+6.5	+4.5

4.3 Equalizer adjustment algorithm

The 4000-mile equalization objective for the L5 system is a very difficult objective to meet. Therefore, great attention has been given to the selection of the Bode network shapes used in the E1 and E2 equalizers and to the determination of the best equalization algorithm so that optimal equalization is achieved from section to section.

From a transmission viewpoint, we can consider the E1 and E2 equalizers as a single equalizer with 28 Bode networks whose frequency response, $EQL(f)$, is given by

$$EQL(f) = \sum_{k=1}^{28} g_k B_k(f) \quad \text{dB}, \tag{2}$$

where f indicates the frequency and $g_k B_k(f)$ represent the gain and frequency response of the kth Bode network, respectively.

If $M(f)$ represents the misalignment to be equalized, the residual error, $E(f)$, will be, after equalization,

$$E(f) = \sum_{k=1}^{28} g_k B_k(f) - M(f) \quad \text{dB}. \tag{3}$$

The purpose of the equalization algorithm is to minimize a certain measure of $E(f)$ in eq. (3) over the frequency range of interest. The ideal condition would be to make $E(f)$ equal to 0 over the entire frequency range by proper selection of $B_k(f)$ and g_k. In reality, however, this is not possible. The L5 equalization strategy is to minimize the mean-squared-error (MSE) function of $E(f)$, which is defined as

$$\text{MSE} = \frac{1}{f_2 - f_1} \int_{f_1}^{f_2} |E(f)|^2 df, \tag{4}$$

where f_1 and f_2 are low and high bounds of the frequency band of interest.

The optimum set of gains, g_k^*, which results in the minimum MSE can be found if the gradient, G_k, of the MSE with respect to each gain, g_k, becomes zero.[10,11] That is,

$$\begin{aligned} G_k = \frac{\partial \text{MSE}}{\partial g_k} &= -2\langle B_k(f), E(f)\rangle \\ &= 2\langle B_k(f), \sum_{j=1}^{28} g_j B_j(f)\rangle - 2\langle B_k(f), M(f)\rangle \\ &= 0 \text{ for all } k = 1, 2, 3, \cdots, 28, \end{aligned} \tag{5}$$

where we define

$$\langle A, B \rangle = \frac{1}{f_2 - f_1} \int_{f_1}^{f_2} A(f)B(f)df.$$

A simultaneous equation expressed by (5) for all $k = 1, 2, \cdots, 28$ may be represented as

$$\mathbf{G} = \mathbf{Bg} - \mathbf{M}, \tag{6}$$

where

$$\mathbf{G} = [G_1, G_2, \cdots, G_{28}]^T, \quad \mathbf{g} = [g_1, g_2, \cdots, g_{28}]^T,$$
$$\mathbf{M} = 2[\langle B_1, M \rangle, \langle B_2, M \rangle, \cdots, \langle B_{28}, M \rangle]^T$$

and

$$\mathbf{B} = 2 \begin{bmatrix} \langle B_1, B_1 \rangle, \langle B_1, B_2 \rangle, \cdots, \langle B_1, B_{28} \rangle \\ \langle B_2, B_1 \rangle, \langle B_2, B_2 \rangle, \cdots, \langle B_2, B_{28} \rangle \\ \vdots \\ \langle B_{28}, B_1 \rangle, \langle B_{28}, B_2 \rangle, \cdots, \langle B_{28}, B_{28} \rangle \end{bmatrix}.$$

The gain vector is

$$\mathbf{g} = \mathbf{B}^{-1}(\mathbf{G} + \mathbf{M}) \tag{7}$$

and the optimum gain setting is obtained if $\mathbf{G}$ in (7) is $\mathbf{0}$, i.e.,

$$\mathbf{g}^* = \mathbf{B}^{-1}\mathbf{M}. \tag{8}$$

The MSE algorithm given in Ref. 10 solves eq. (7) with $\mathbf{G} = \mathbf{0}$ by the steepest descent method. This may be readily implemented in an automatic equalizer control circuit, but not in the manually adjusted equalizers. The steepest descent algorithm requires simultaneous adjustment of all the gain settings. An equalizer adjustment algorithm based on the Gauss-Seidel iterative method discussed in Ref. 11 is suitable and hence adapted for the manual E1 and E2 equalizer adjustment. This algorithm calls for the following equalizer gain adjustment procedure:

(*i*) Adjust the first gain, g_1, until its corresponding gradient $G_1 = 0$.

(*ii*) Repeat for the second gain, g_2, through the last one, g_{28}, until corresponding gradients become zero, thus completing one iteration.

(*iii*) Repeat the above steps.

As the number of iterations is increased, the equalizer settings approach the optimum point described by eq. (8). A detailed mathematical analysis of this procedure is presented in Ref. 11.

So far, it has been assumed that the gradient G_k can be continuously calculated during the adjustment of the kth Bode network gain, g_k.

One approach is to compute the gradient by eq. (5). The gradient is obtained by cross-correlating the equalizer shapes $B_k(f)$ with error $E(f)$. This procedure requires continuous error information over the frequency range of interest and results in a rather complex hardware realization.

It is shown in Ref. 10, however, that the gradient of the MSE with respect to a particular network gain can be very closely approximated by the following simple relationship:

$$G_k = \sum_{i=1}^{3} B_k(f_{ki})E(f_{ki}), \tag{9}$$

where $E(f)$ is the channel residual error defined in eq. (3), f_{k2} is the center frequency of the Bode network $B_k(f)$, and f_{k1} and f_{k3} are lower and upper side frequencies of $B_k(f)$ such that

$$B_k(f_{k1}) = B_k(f_{k3}) = \tfrac{1}{2}B_k(f_{k2}).$$

If $B_k(f_{k2})$ is normalized and its value is 1, eq. (9) becomes simply

$$G_k = \tfrac{1}{2}E(f_{k1}) + E(f_{k2}) + \tfrac{1}{2}E(f_{k3}). \tag{10}$$

In other words, by sampling the channel error at three frequency points and by properly weighting these errors, we may approximate the necessary gradient information. This approach is called the simplified MSE algorithm.

In Ref. 11, it is shown that a further approximation has resulted in a very simple way of computing the gradient:

$$G_k = 2E(f_{k2}). \tag{11}$$

That is, the gradient is approximated by sampling the channel error at a single frequency (usually the center frequency of the Bode network). An equalizer adjustment method with the gradient obtained as in eq. (11) has been referred to as the zero forcing (ZF) algorithm, and has been used, for example, in the L4 line equalization. When the set of shapes B_k are near-orthogonal, the ZF algorithm results in a rapid convergence to the optimum gain settings.

Extensive computer simulation and field testing have been conducted to compare the effectiveness of the three adjustment algorithms in which the gradients have been calculated according to eqs. (5), (10), and (11). The results are shown in Fig. 19, which indicates that there is considerable improvement in the simplified MSE algorithm over the ZF algorithm. However, little improvement is achieved when the gradients are calculated according to the more complex relationship

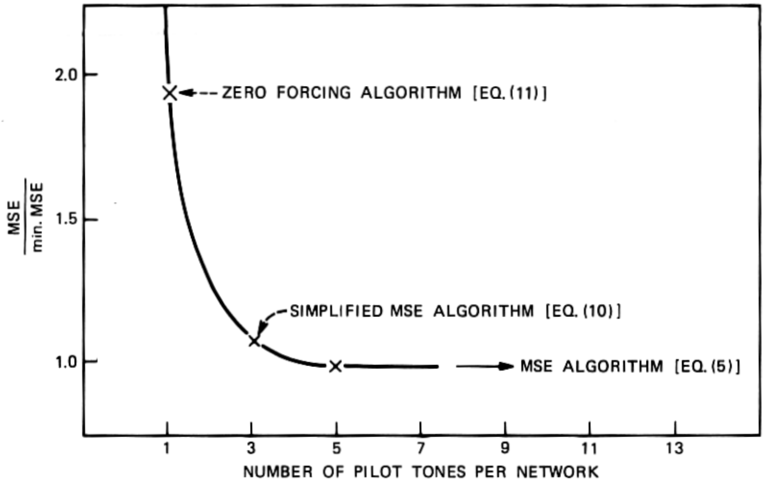

Fig. 19—MSE values by different algorithms.

in (5). In the L5 equalization plan, the simplified MSE algorithm has been implemented with the aid of a unit called the equalizer adjustment unit (EAU).

4.4 Equalizer adjustment unit and equalization process

The EAU was developed to aid in the adjustment of the E1 and E2 equalizers according to the iterative procedure described in the previous section with the gradient information computed by either the simplified MSE or ZF algorithm. The EAU is a hardwired special-purpose computer containing a preprogrammed memory and an arithmetic unit.

In the equalization process, the EAU is used in conjunction with the 90-type transmission measuring set composed of the 90G oscillator, 90H detector, and 90F digital control unit,[12] which is located in every L5 main station.

For equalization, two separate EAU's and transmission sets are required, one at the transmitting station and one at the receiving station (see Fig. 20). During equalization, each EAU is set by its operator for a particular Bode network, say B_k, and the EAU outputs the appropriate control information to the 90 set for that network. At the transmitting station, the command portion of the EAU instructs the 90G to generate specified levels and proper frequencies corresponding to the network to be adjusted. In the ZF mode, the EAU causes the 90G to generate only frequency f_{k2}, which corresponds to

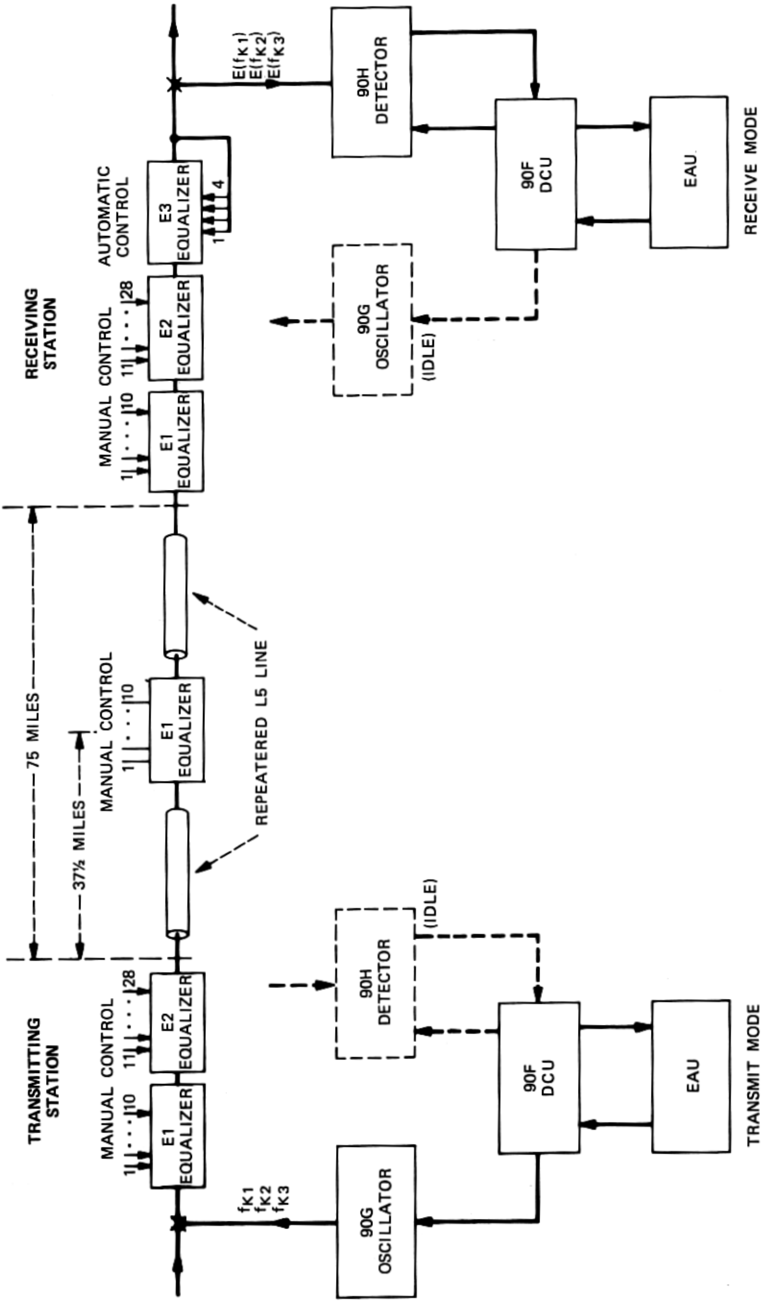

Fig. 20—Equalization plan.

the center frequency of B_k. If the MSE mode is selected at the EAU, the 90G will generate sequentially frequencies f_{k1}, f_{k2}, and f_{k3} as defined in eq. (10). These frequencies are then transmitted on the L5 line.

At the receiving location, the EAU initially establishes synchronization between the transmitting 90G and the receiving 90H. The 90H then receives these signals, measures the received level, and transmits the resulting level information to the arithmetic portion of the EAU. There it is processed and the resultant gradient information is displayed to the operator in numeric form. The displayed gradient is calculated by either eq. (9) or (11), depending on whether the MSE or ZF mode, respectively, has been selected by a front panel switch located on the EAU.

In the equalization procedure, since the gradient information is displayed only at the receiving end of the line, the adjustment is under the control of the operator at the receiving station. The receiving operator relays the correct gain setting for each equalizer bump to the operator at the transmitting station and to the craftsperson at the manhole equalizing repeater location. To achieve the proper pre- and postequalization during the adjustment of the E1 equalizer, one-half the error is compensated for by the equalizer in the manhole and one-quarter of the error is compensated for in both the transmitting and receiving main-station equalizers. In the case of the E2 equalizer, one-half the correction is inserted at both the transmitting and receiving main stations. The ultimate objective is to adjust all the equalizer shapes so that the resultant gradient, as displayed on the EAU, is zero for all shapes.

In the equalization procedure, the ZF algorithm is first used because of its rapid convergence to the near-optimum gain settings. The MSE algorithm is then used to "fine tune" the equalization. The iterative procedure is as follows:

(*i*) Select the ZF mode of gradient calculation.
(*ii*) Select E1 equalizer, shape B_1. Adjust the equalizers (one-half the correction in the manhole equalizer, one-quarter in both transmitting and receiving main stations) until the gradient of the error is less than 0.05 dB.
(*iii*) Repeat for E1 shapes B_2 through B_{10}.
(*iv*) Select E2 equalizer, shape B_{11}. Adjust the equalizers (one-half correction in both transmitting and receiving main stations) until the gradient of the error is less than 0.05 dB.

(*v*) Repeat (*iv*) for E2 shapes B_{12} through B_{28}.
(*vi*) Select the MSE mode of gradient calculation.
(*vii*) Iterate steps (*ii*) through (*v*) until the gradient is within ± 0.05 dB of zero for all shapes.

In practice, convergence of the iteration is quite rapid so that usually two or three iterations are sufficient.

V. REGULATING REPEATER

5.1 Regulating repeater function

Referring once again to eq. (1), the second bracketed term describes the change in cable attenuation as a function of the cable temperature, T. It is important to note that this change in attenuation varies with frequency in a $\sqrt{f}$ manner. The basic function of the regulating repeaters is to equalize this dynamic attenuation.

In both the L4 and L5 systems, the coaxial cable is buried at a depth of four feet. The change in cable loss resulting from a $\pm 20°$F temperature change, for power-feed spans approaching 75 miles in length, is ± 52 dB at 66 MHz. This is the magnitude of the equalizing function of the regulating repeaters at the high end of the L5 frequency spectrum.

The strategy in providing compensation for this seasonal dynamic change in loss is focused at minimizing the signal-to-noise impairment resulting from misalignment caused by the temperature effect. In the L5 system, the distance between regulating repeaters is a maximum of seven miles. With this bound established, the misalignment resulting from cable temperature changes is reduced from ± 52 dB to ± 4.8 dB.

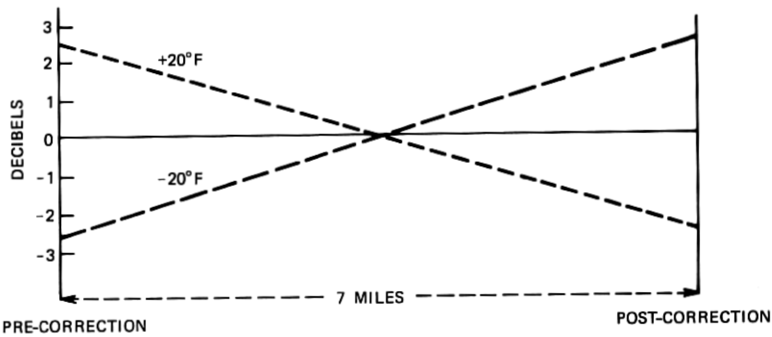

Fig. 21—Regulating section—66-MHz misalignment at the regulating repeater output and input owing to cable temperature variation.

When both pre- and post-correction are provided, the misalignment is further reduced to ± 2.4 dB as shown in Fig. 21.

A second effect that the regulating repeater must compensate for is not directly related to temperature, but more to distance. Since the basic repeater is a fixed gain amplifier designed to equalize the loss of one mile of cable at 55°F, adjustments are made in the nominal spacing of the repeater manholes when cable is placed in areas having other than a mean annual temperature of 55°. In warmer areas, the manholes are placed at intervals shorter than 1 mile; in colder areas, placement intervals are greater than 1 mile. A 5-degree error in mean annual temperature estimation for system design results in 2158 feet of cumulative error in manhole placement over 75 miles. This amounts to 13.1 dB of misalignment at 66 MHz or, as in the case of the L5 system, 1.2 dB in a 7-mile regulating section. The loss relationship as defined in (1) is predominantly $\sqrt{f}$.

There is a third source of misalignment. One inevitable situation that plagues those who engineer cable systems is the inability to place a manhole at the locations dictated by the nominal spacing. Right-of-way procurement, convenient access, terrain considerations, and densely populated and urban areas all contribute to a deviation from nominal spacing. The accumulation of these deviations within a regulating section results in additional misalignment that is accommodated by the line-build-out networks in 0.1-mile increments, and the remainder is equalized by the closed-loop regulating action of the postregulator described in the next section. Therefore, equalization of cumulative spacing deviations of ± 0.05 mile, equivalent to ± 1.6 dB at 66 MHz, is another objective of the regulating repeater.

Other sources of misalignment, such as the deviation in basic repeater gain from the nominal gain, the uncertainty in placed cable lengths, and others, are estimated to be ± 0.4 dB. Table II lists the misalignments and establishes the total range required of both the post- and preregulator circuits at 66 MHz for 7 miles.

Table II — Misalignments allocated to the regulating repeater

Cause	Magnitude
Change in cable loss caused by temperature variation (± 20°F)	± 4.8 dB
Cumulative length error (± 0.038 mi)	± 1.2 dB
Cumulative spacing deviation (± 0.05 mi)	± 1.6 dB
Other	± 0.4 dB
Total	± 8.0 dB

The preregulator automatically accommodates for ±2.4 dB of the cable loss variation resulting from temperature changes, and the postregulator must accommodate the remaining ±5.6 dB. This effect is illustrated in Fig. 22.

This suggests a range requirement of 11.2 dB for the postregulator. However, because of circuit realization limitations, only 8.7 dB was attainable in either a post- or preregulator network while still meeting a broadband objective of less than ±0.015 dB per dB of tracking error. It was found that line acceptance procedures overcome this limitation if the uncertainties and deviations accumulate in such a way that the range of the regulating repeater is exceeded. These procedures locate the excess variations which are then compensated for with either reassignment of line-build-out networks or readjustment of preregulators.

The total function of the regulating repeater, therefore, is:

(*i*) To equalize the dynamic temperature-dependent term of the cable equation in a pre- and postequalization strategy.
(*ii*) To accommodate automatically the uncertainties and cumulative spacing deviation in a postequalization strategy.

This total function is aimed at minimizing the signal-to-noise impairment by maintaining the distributed misalignment at small magnitudes.

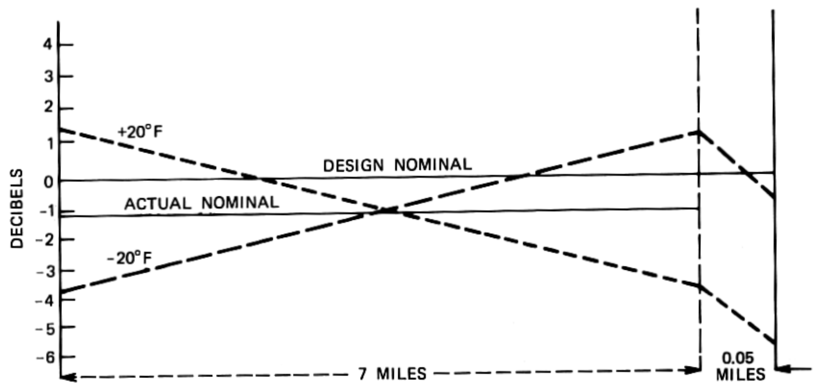

Fig. 22—Regulating section—66-MHz misalignment at the regulating repeater output and input resulting from cable temperature variation and other sources of misalignment.

5.2 Regulating repeater circuits

The regulating repeater consists of those circuits that comprise a basic repeater plus additional circuits to perform the pre- and post-regulating functions. These additional circuits are located between the preamplifier and power amplifier as shown in Fig. 23.

It has been established that the misalignments allocated to the regulating repeater are predominantly losses or gains that vary with frequency in a $\sqrt{f}$ manner because of temperature and length. Therefore, the equalization networks must also vary as $\sqrt{f}$. The misalignment allocated to the preregulator is one-half the variation resulting from temperature. Control of the preregulator is established by sensing earth temperature in the vicinity of the manhole. To establish proper tracking of the cable temperature, the sensor must be placed in an earth environment similar to that in which the cable is buried. The depth of the sensor is particularly important. Errors in depth significantly affect the amplitude and delay of earth temperature variations with respect to the cable.

Control of the postregulator is established by sensing the level of a single frequency, designated the temperature pilot, which is applied at the transmitting end of the cable. This method of control has the advantage of automatically correcting for all misalignments of this pilot. Therefore, the remaining half of the temperature effect on cable, cumulative spacing deviations, and uncertainties in cable length and design temperature are equalized in the postregulator. However, pilot control may be troublesome if the losses incurred are not $\sqrt{f}$, since the correct equalization shape will not be provided. In the L5 system, this difficulty is avoided by the proper selection of the pilot frequency and the attention given to the basic repeaters and to the fixed and adjustable equalizers at this frequency. Although maximum sensitivity to cable temperatures occur at the high end of the L5 spectrum, the pilot was chosen to be at 42.880 MHz. The reasons for this choice are

(*i*) The temperature coefficient of the basic repeater (not $\sqrt{f}$) could be made to approach 0 at this frequency, but is a maximum at higher frequencies.

(*ii*) Return loss is better controlled and therefore interactive effects are minimized.

(*iii*) This frequency is in the guard band between the second and third jumbogroups.

(*iv*) This frequency slot is void of multiplex carrier leaks.

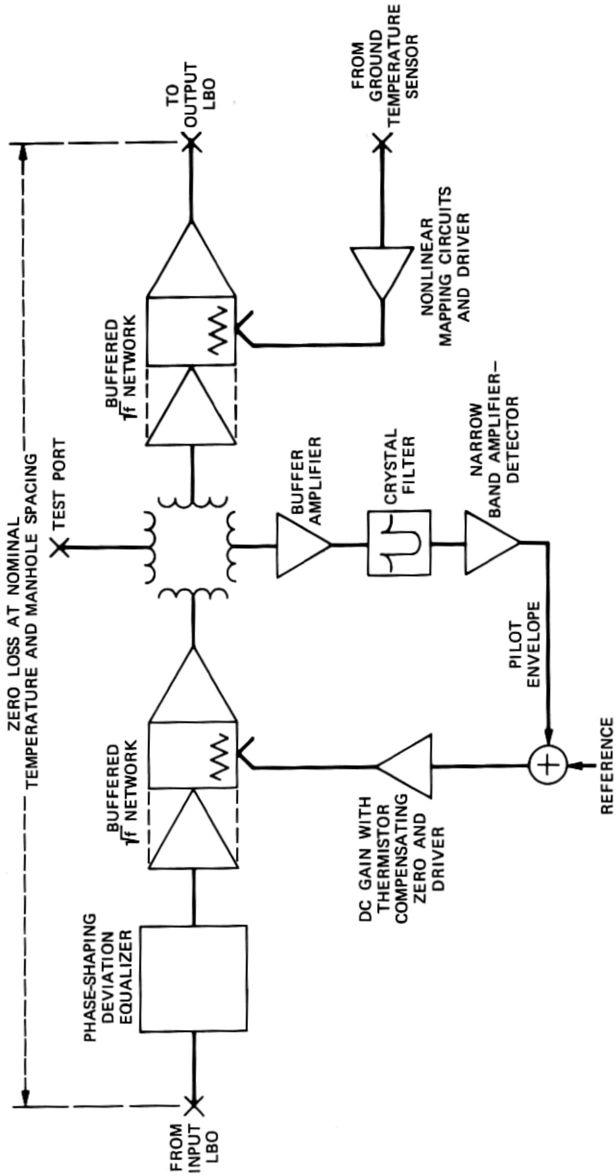

Fig. 23—Additional circuits that perform the pre- and postregulating functions.

Each pre- or postregulator contains identical Bode equalizers ($\sqrt{f}$ networks[9]) with loss adjustment controlled by varying a single resistive element. This element is an indirectly heated thermistor with the heater winding electrically isolated from the thermally controlled resistor.

Gain is provided in each regulator to compensate for the loss in the Bode equalizer and the associated networks by single-stage thin-film hybrid integrated-circuit amplifiers. Two amplifiers and the Bode equalizer are packaged to form the $\sqrt{f}$ network. The amplifiers also provide the proper terminating impedances for both ports of the equalizer to achieve an accuracy that is better than 1 percent in matching the loss variation which is $\sqrt{f}$. The type of transistor used in the basic repeater circuits is also used in the single-stage amplifier to achieve the noise figure and linearity requirements.

One critical aspect in the design of the preregulator control circuitry is to accurately map the nonlinear resistance-versus-temperature function of the buried ground temperature sensor into a linear loss (in dB) function. In addition, provision had to be made to control the sensitivity of the function since regulating sections can vary significantly in length, and to adjust for the mean annual temperatures since the same sensor is placed in all climatic areas. The response of the temperature sensor is shown in Fig. 24. The range expected is $\pm 20°F$

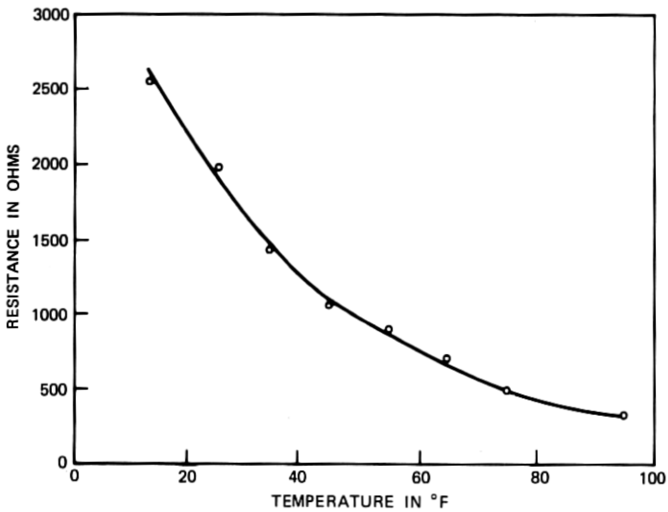

Fig. 24—Ground-sensing thermistor—resistance vs temperature.

about a mean annual temperature which in itself can vary $\pm 20°$ from 55°F.

The nonlinear mapping was accomplished through the use of operational amplifiers. With the use of negative feedback and scaling, the control circuit provides a temperature tracking function that is accurate to better than 2°F. The mean annual adjustment is made accessible so that readjustments may be made safely in the manhole when the repeater is installed and powered, should line acceptance procedures indicate that the postregulator range is being exceeded.

In the postregulator, the temperature pilot is selected from the message band with a crystal filter, amplified, and converted to a direct current in a peak detector. Comparison of this direct current signal to a reference results in an error signal that is amplified and applied to the heater of the thermistor in the Bode equalizing network. The heater sets the thermistor resistance in the Bode network and the pilot level is maintained constant at the output for changes in level at the input.

The goal in this control circuitry is to achieve a reasonable response time while maintaining a well-damped, nonenhanced system of many postregulators in tandem.

A system of tandem regulators may be modeled as shown in Fig. 25. In this figure,

E_{pi} = the envelope that might exist on the pilot at the input of the system

E_{po} = the envelope on the pilot at the end of a system of n regulators

E_{mo} = the envelope that results in the message band

$G(s)$ = the control loop transfer function of a single regulator, where s is a Laplace transformed variable.

The normal expected response of this system is to have E_{pi} vary seasonally while tracking each cable section loss change that results from the cable temperature change. This variation is extremely slow (one cycle per year) and, since $G(s)$ is large for such a slow variation, the envelope gain of the pilot is nearly zero while the message envelope is properly tracking ($A = \sqrt{f_m/f_p}$) the pilot envelope to maintain the levels constant at the output of each regulator. Therefore, each regulator compensates for its preceding section of cable.

For very fast variations in the pilot envelope, the envelope gain to the pilot is unity, and the message envelope gain is nearly zero. Message channels, therefore, are not affected by the faster perturbations.

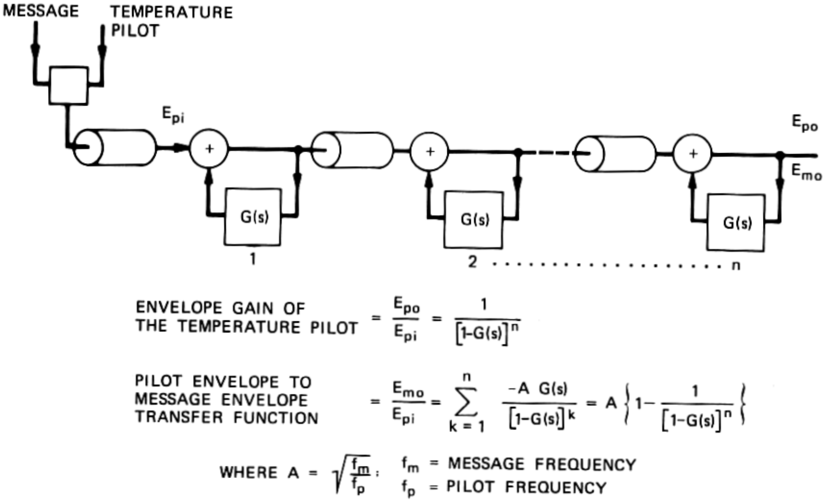

Fig. 25—System of tandem postregulators.

When $G(s)$ has a single pole, no gain enhancement ($E_{po}/E_{pi} > 1$) will occur, and minimum overshoot to transients will result. When $G(s)$ has more than a single pole, gain enhancement will occur where $G(s)$ approaches unity, and the system will not be critically damped to transients.

In the postregulator, the loop response contains several poles. These poles are a result of the indirectly heated thermistor. A plot of the frequency response of the loop gain is shown in Fig. 26.

To restrict the gain enhancement to a tolerable magnitude, the response is compensated for by using an operational amplifier to provide a pole cancelling zero in the vicinity of 2 Hz. This results in an enhancement response as shown in Figs. 27 and 28 where computed and measured data are compared and in a transient response as shown in Fig. 29.

One final aspect of the circuitry pertains to a transfer arrangement in the event of pilot failure. The loss of the temperature pilot initiates a switch to the spare line. This pilot loss could also cause the envelope-controlled regulators to go to an extreme gain condition. To prevent this high gain from occurring, heater control of the postregulator is transferred from envelope control to preregulator control. Under these circumstances, those misalignments allocated to the postregulator are no longer equalized. However, in an L5 switching section (150

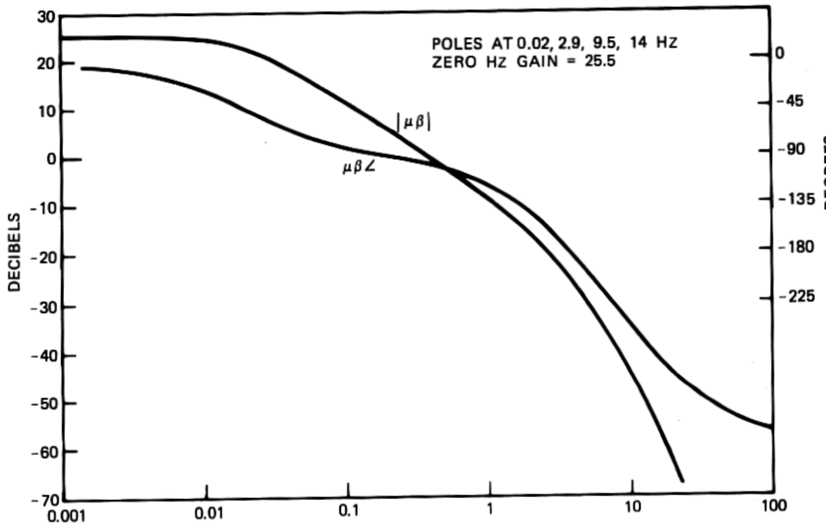

Fig. 26—Postregulator $\mu\beta$ response equivalent to $G(s)$ in the regulator model.

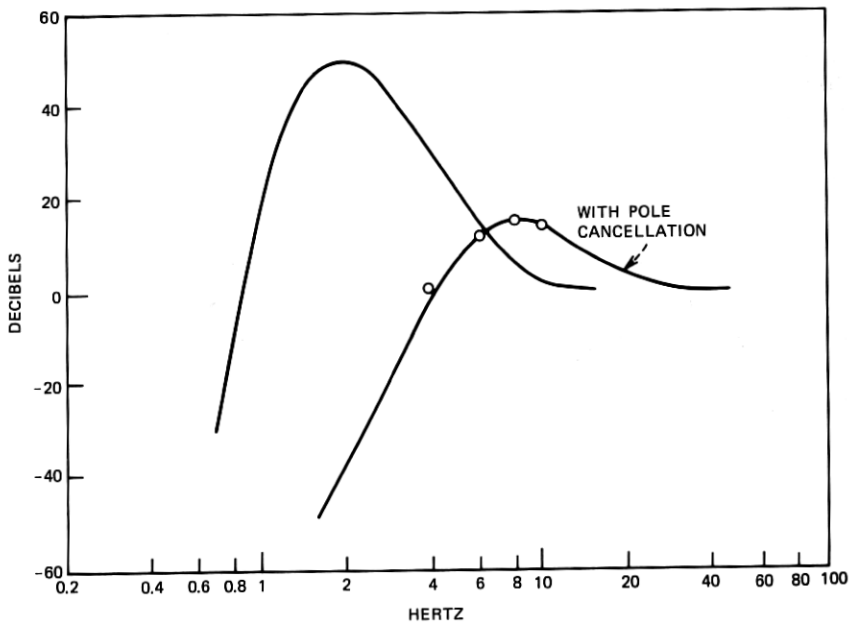

Fig. 27—Computed pilot envelope response for 80 regulators in tandem with and without thermistor pole cancellation. Measured data points are compared.

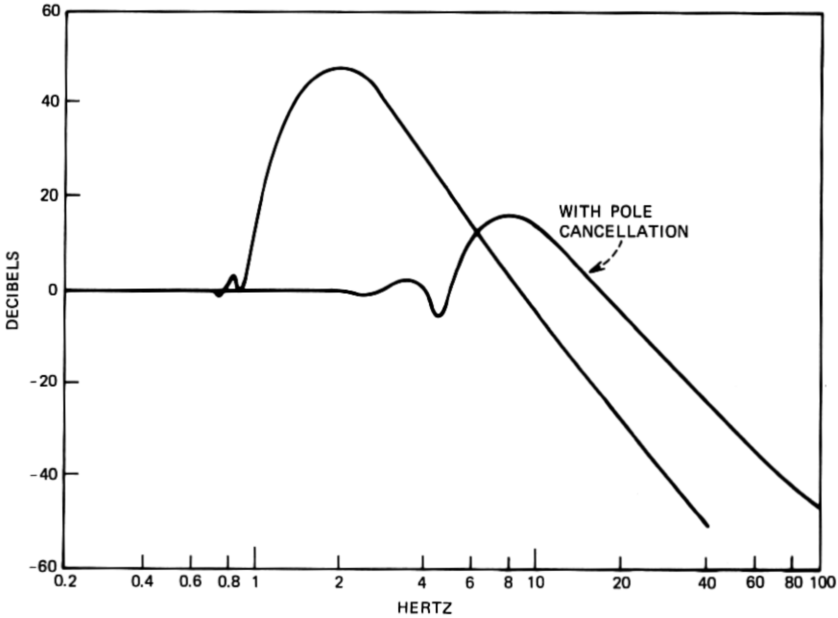

Fig. 28—Computed channel envelope response for 80 regulators in tandem with and without thermistor pole cancellation.

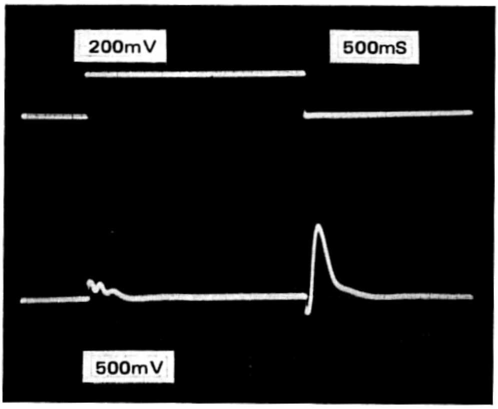

Fig. 29—Transient response of 80 regulators in tandem. Upper trace is the input excitation of 0.8 dB.

miles), the misalignments owing to cumulative spacing deviation and uncertainties tend to be both positive and negative, and hence there is some cancellation. The net result is that gain is maintained near normal. Also, regulator response time is fast so recovery from typical no-pilot misalignments of 6 dB is complete in less than 5 seconds for a switching section containing 25 regulators.

An operational amplifier is used to assist in accomplishing an accurate transfer. Heater current to the preregulator and pilot envelope are monitored. When the pilot level drops below a prescribed value, heater current which is identical to that flowing in the preregulator is caused to flow in the heater of the postregulator, and the two Bode networks track the ground temperature sensor.

5.3 Regulating repeater noise performance

The networks in the transmission path of the regulating repeater contribute to the noise and linear performance of the repeater. These parameters are listed and compared to the same parameters of the basic repeater in Table III.

VI. THE E3 DYNAMIC EQUALIZER

The final level of equalization in the L5 system is performed by an automatic pilot controlled equalizer which equalizes for a number of dynamic effects. The primary effect is that associated with the small but significant change in the gain of the L5 line repeaters because of seasonal temperature changes. Repeaters located in manholes are subject to ambient temperature changes roughly equivalent to the ±20°F temperature change of the coaxial cable itself. Line repeaters having an average temperature coefficient of less than 0.002 dB per degree Fahrenheit at 66 MHz introduce less than 3-dB change in line gain over a 75-mile power-feed section. The variation of this deviation with frequency is predominantly non-$\sqrt{f}$ and is therefore not compensated for by the regulating repeaters.

Table III

	Basic		Regulating	
	Low Freq. (dB)	High Freq. (dB)	Low Freq. (dB)	High Freq. (dB)
NF	8.3	5.5	13.5	8.2
M_{2E}	−108	−74	−90	−65
M_{3E}	−135	−110.5	−115	−104.5

In addition, as discussed in the previous section, a tracking error is associated with the inability of the regulating repeaters to accurately match the change in loss of the cable as a function of frequency and temperature. Over a 75-mile power-feed section, deviations of up to 1.6 dB at 66 MHz can be expected.

While seasonal temperature variations have the main effect on line gain change, there are other second-order, longer term effects such as component drift, aging, and increases in the manhole ambient temperature as the number of equipped coaxial lines in a route is increased. All these taken together represent the total dynamic deviation of the L5 line and are equalized by the E3 dynamic equalizer.

The realization of the E3 equalizer has been a function of a number of system considerations. In previous sections, it was pointed out that the static deviation, being a result of statistical manufacturing deviations of many components, was impossible to predict *a priori*. Thus the E1 and E2 equalizers are very flexible and have many degrees of freedom to match a wide variety of shapes.

On the other hand, we expect the dynamic deviation to be much better behaved. In previous L-carrier systems, the gain of the line repeater as a function of the repeater temperature was a broad function of frequency. Predictions based upon gain sensitivity analysis indicated a similar behavior for the L5 repeatered line. We would expect then that an equalizer designed to compensate for this effect would also be a broad function of frequency with correspondingly fewer degrees of freedom. Furthermore, this function could be measured *a priori* by measuring the gain deviation of a sample of repeaters for various ambient temperatures.

While the dynamic misalignment of the L5 line is a slowly varying function of frequency, the overall equalization objective of the L5 line dictates that automatic equalization be employed rather than manual equalization, as in the case of E1 and E2 equalizers. While there are many ways we could conceive of providing automatic equalization, the use of pilot tones in the composite message signal and independent of the message signal itself allows a relatively simple and reliable method of providing such equalization. Therefore, pilot frequencies have been included above and below the message band as well as between jumbogroups to provide for dynamic equalization. The pilot between jumbogroup 2 and 3 is the same pilot used for temperature regulation.

All these considerations have led to the E3 dynamic equalizer with four degrees of freedom, each degree of freedom being automatically controlled by a line pilot.

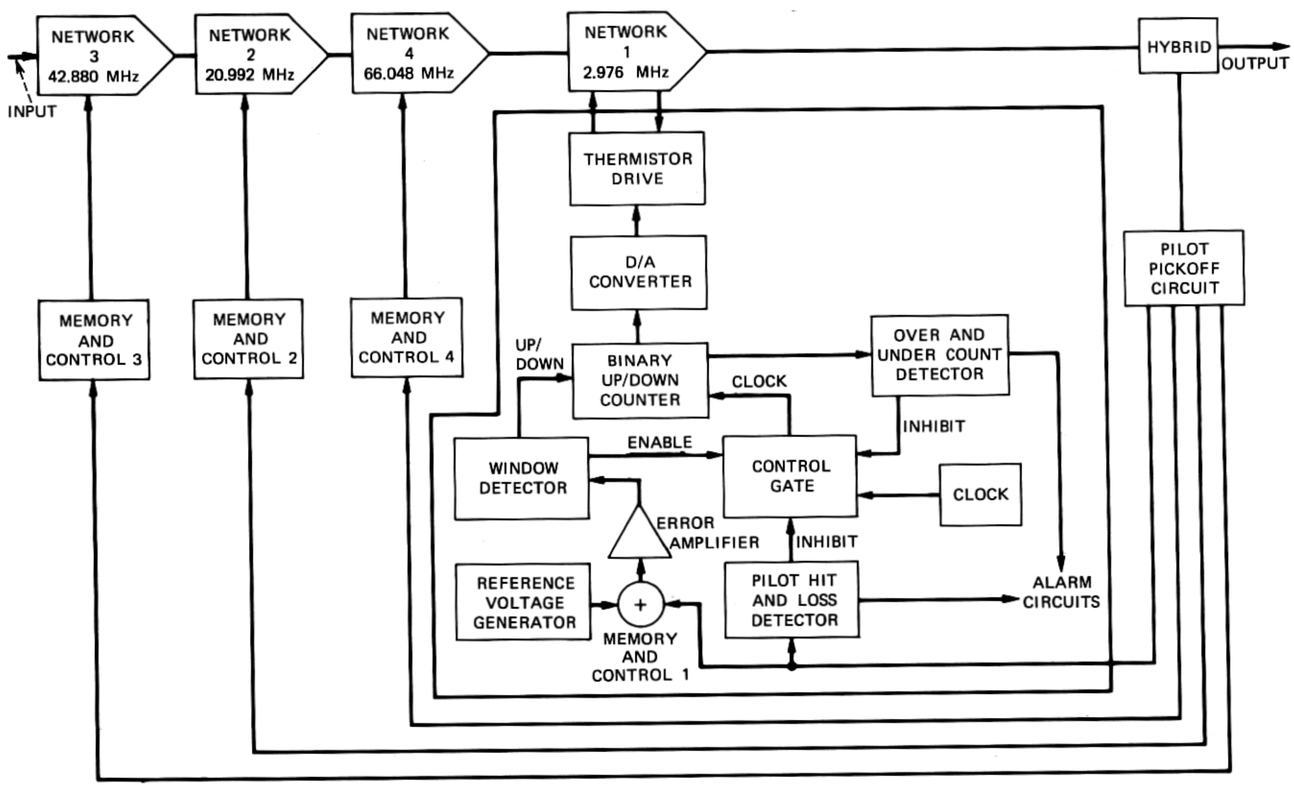

Fig. 30—E3 dynamic equalizer.

Unlike all other equalization in the L5 system, the E3 dynamic equalization is provided only on a postequalization basis. There are two reasons for this: In the first place, the expected misalignment over a 75-mile power feed span is about 3 dB at 66 MHz. The noise penalty associated with postequalization only, under these conditions, is on the order of 1 dB and has been included in the system margins. In addition, any scheme to provide both pre- and postequalization on a dynamic basis for all four pilots would be considerably more expensive and is not economically justifiable. Thus, the E3 dynamic equalizer is located in main stations at the receive end of each L5 power-feed section.

6.1 E3 equalizer high-frequency networks

Figure 30 is a block diagram of the E3 dynamic equalizer. Basically, it consists of four high-frequency networks through which the L5 line signal passes. Each network has a variable gain shape such that the combined frequency characteristic of all four networks closely matches the expected residual dynamic misalignment of the L5 line. Three of these shapes are Bode bumps, while the fourth shape is a flat function of frequency (Fig. 31).

In Fig. 30, the first network is a series Bode network imbedded between two flat-gain amplifiers. The amplifiers provide an overall gain of 6.55 dB and provide good input and output return loss. Im-

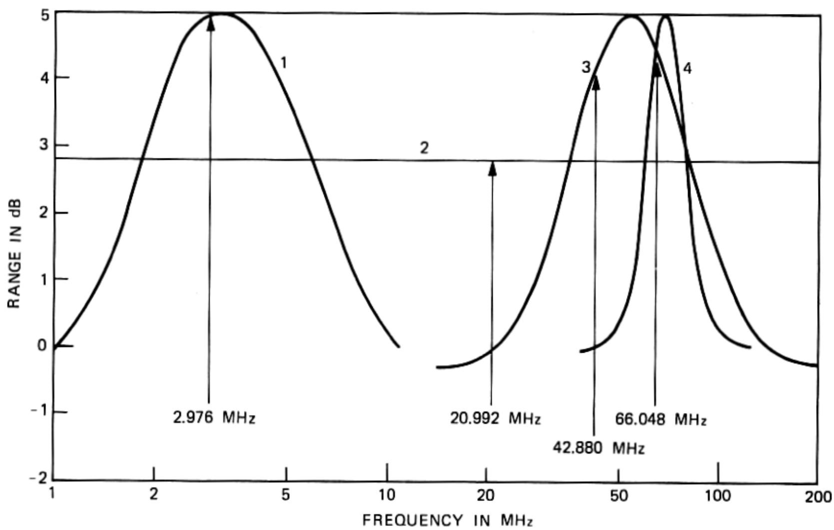

Fig. 31—E3 equalizer shapes.

bedded in the Bode network is a thermistor whose resistance is controlled by the 42.880-MHz temperature pilot at the E3 equalizer output through the control circuit feedback path. This resistance in turn controls the gain of the Bode network such that the level of the 42.880-MHz pilot at the output of the equalizer is held at a fixed reference level.

The second network is a broadband variolosser that provides a flat gain or loss around a nominal loss of 12 dB. The amount of gain or loss is controlled by a thermistor in the variolosser whose resistance is controlled by the level of the 20.992-MHz pilot at the output of the equalizer.

The third network is a series Bode network similar to the first network. In this network, the gain of the Bode network is determined by the level of the 66.048-MHz pilot at the output of the equalizer.

The fourth network is a broadband feedback amplifier of nominal 14.0-dB gain with a Bode network in its feedback path. This network also contains a thermistor whose resistance is controlled by the level of the 2.976-MHz pilot at the equalizer output.

6.2 Control circuits

The E3 equalizer control circuits contain a pilot pick-off circuit where the four E3 line pilots are separated from the message signal and converted to four dc voltage levels, each proportional to the absolute power level of the respective pilot. The remainder of the control circuits are digital circuits. The heart of each digital circuit is a binary up-down counter coupled with a digital-to-analog converter that acts as a digital integrator in the feedback loop. The system works in the following fashion: If the pilot level at the output of the equalizer deviates more than about 0.05 dB from nominal, it is detected by a change in the dc pilot voltage. Under this condition, the up-down counter is allowed to count. This count is converted by the digital-to-analog converter to a dc current, which is then applied to the thermistor in the corresponding network. This causes the network gain to change in a direction to restore the output pilot level to its correct value. When the pilot returns to within ± 0.05 dB of nominal, the counter is stopped and the gain of the network is held at that level.

There are several advantages in using digital circuits in the feedback loop when space and power consumption are not critical.

(*i*) With a digital integrator within the control loop, an equivalently higher loop gain can be obtained than with an analog integrator. The steady-state accuracy of the control loop is determined primarily by the size of the quantizing steps, the tracking between the digital-to-analog converter, and the network control characteristic (input voltage versus network gain in dB).

(*ii*) If an abrupt change in pilot level or loss of pilot occurs because of an abnormal line condition, the digital memory holds the equalizing network control fixed.

(*iii*) The midrange gain of each equalizing network is precisely defined by the state of the up-down binary counter so that the network may be set at midrange with a local key or by remote command.

(*iv*) The control loop dynamic response can be controlled by the rate at which pulses are clocked into the up-down counter.

(*v*) The binary counter state can be monitored to indicate equalizer gain and provide for control and alarm functions such as locking the gain at any given state and giving a warning of end of equalizing range.

6.3 Dynamic behavior of the E3 equalizers[13]

If $M(f, t)$ represents the time-varying misalignment to be equalized, the residual error after equalization will be similar to eq. (3) and can be expressed by

$$E(f, t) = \sum_{k=1}^{4} g_k(t)B_k(f) - M(f, t), \tag{12}$$

where f and t indicate the frequency and time, g_k and B_k are the gain and input-output relationship of the kth network, respectively, and there are four adjustable networks in the equalizer.

The E3 equalizer is adjusted continuously in-service by sampling channel misalignments at four pilot frequencies. In this way, four pilot signals generate the four equalized channel errors. The network gains are adjusted until the four errors become zero.

The block diagram shown in Fig. 30 can be represented by the functional block diagram shown in Fig. 32. Figure 32 indicates that the E3 equalizer is a multivariable system whose input and output relationship can be expressed by the following equation.

From eq. (12),

$$\mathbf{E} = \mathbf{Bg} - \mathbf{M}, \tag{13}$$

where $\mathbf{E} = [E(t, f_1), E(t, f_2), E(t, f_3), E(t, f_4)]^T$, f_1, f_2, f_3, and f_4 are the

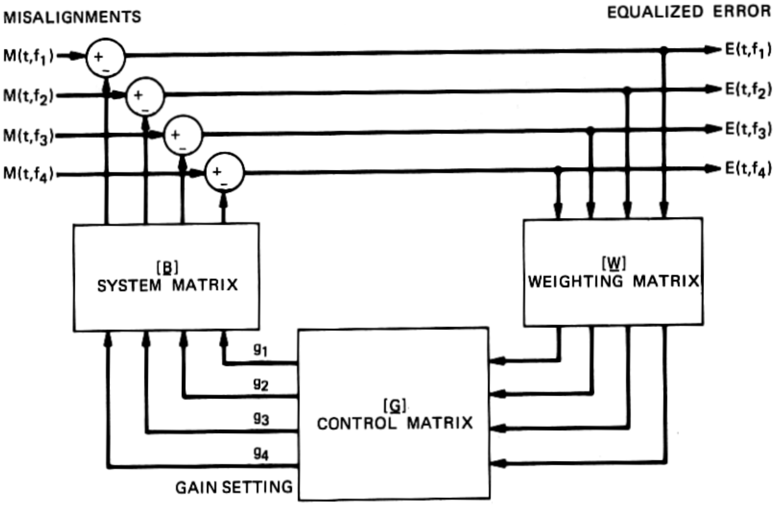

Fig. 32—E3 equalizer functional block diagram.

four pilot frequencies, $\mathbf{M} = [M(t, f_1),\ M(t, f_2),\ M(t, f_3),\ M(t, f_4)]^T$, $g = [g_1(t),\ g_2(t),\ g_3(t),\ g_4(t)]^T$, and

$$\mathbf{B} = \begin{bmatrix} B_1(f_1), & \cdots, & B_4(f_1) \\ \vdots & & \\ B_1(f_4), & \cdots, & B_4(f_4) \end{bmatrix}.$$

Referring to Fig. 32, the gain $\mathbf{g}$ is obtained in the feedback loop and expressed by

$$\mathbf{g} = \mathbf{GWE}, \tag{14}$$

where $\mathbf{G}$ is a 4×4 control matrix, and $\mathbf{W}$ is a 4×4 weighting matrix. $\mathbf{G}$ is an operator and determines the dynamic behavior of the four feedback control loops and is expressed approximately by

$$\frac{K}{s(sT + 1)}\,\mathbf{I},$$

where K and T are the constants and $\mathbf{I}$ is a 4×4 unity matrix. From eqs. (13) and (14), the input-output relationship of the equalizer becomes

$$\mathbf{E} = [\mathbf{I} + \mathbf{BGW}]^{-1}\mathbf{M}. \tag{15}$$

The dynamic behavior and system stability are determined by eq. (15).

6.4 Response of tandem-connected equalizers

All the main and power-feed stations along the L5 coaxial line are equipped with E3 equalizers. Hence, any disturbance leaving one equalizer will be propagated to the following equalizers whenever those equalizers are controlled by the same pilot signals. If pilot signals are blocked and reinserted every Nth station, then up to N equalizers are effectively connected in tandem, and when the individual feedback loop is designed, it is necessary to consider the effects on the subsequent $N - 1$ equalizers.

The dynamic behavior of the E3 equalizer is mainly determined by $\mathbf{B}$ and $\mathbf{G}$ in eqs. (13) and (14), respectively. As shown in eq. (13), the matrix $\mathbf{B}$ is determined by the network shapes in E3 and pilot frequencies. When $\mathbf{B}$ and $\mathbf{G}$ are finally designed, a further improvement may be obtained by a suitable choice of the weighting matrix $\mathbf{W}$ shown in eq. (14).

In reality, the control loop transfer function, $\mathbf{G}$, includes nonlinear elements (e.g., thermistor), and the frequency domain approach to the analysis and synthesis of the control loop to satisfy the transient behavior becomes less accurate. A digital computer simulation in the time domain was developed to predict the behavior of the N-tandem-connected equalizers when the transfer function of the single control loop, G, is known. The computer results were used to modify the control loop transfer function to satisfy the system requirements of the N-tandem-connected equalizers.

Figure 33 shows a transient response of four E3 equalizers connected in tandem when input pilots are step-disturbed by 2 dB. Figure 34 is

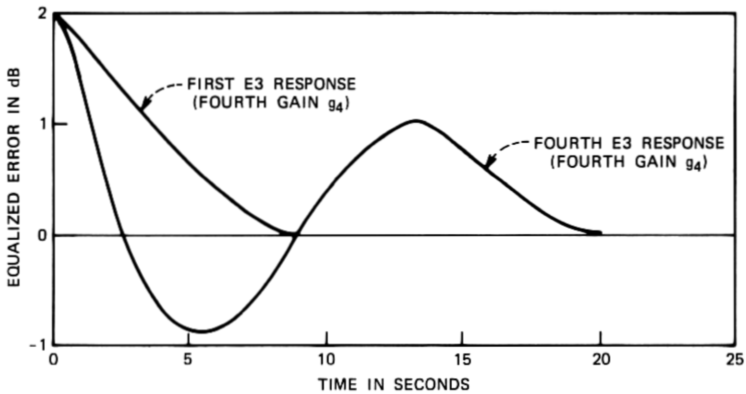

Fig. 33—Transient response of four tandem connected E3 equalizers.

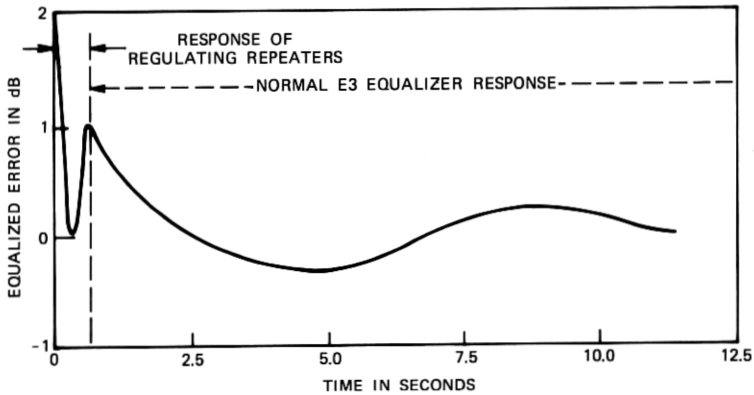

Fig. 34—Transient response of four power-feed main stations (300 miles).

similar to Fig. 33, but includes the transient effects of 40 regulating repeaters within four power-feed sections (about 250 miles). These results were measured in the first L5 installation.

VII. SUMMARY AND CONCLUSION

The concept of a hierarchy of repeaters in which the basic repeater is the fundamental building block was originated in the L4 system.[4] Experience with the L4 system has proven this concept to be sound, and the L5 repeaters are similarly based. Design emphasis was focused on the realization of an ultralinear low-noise repeater with a frequency response that is consistently reproducible in manufacture. This

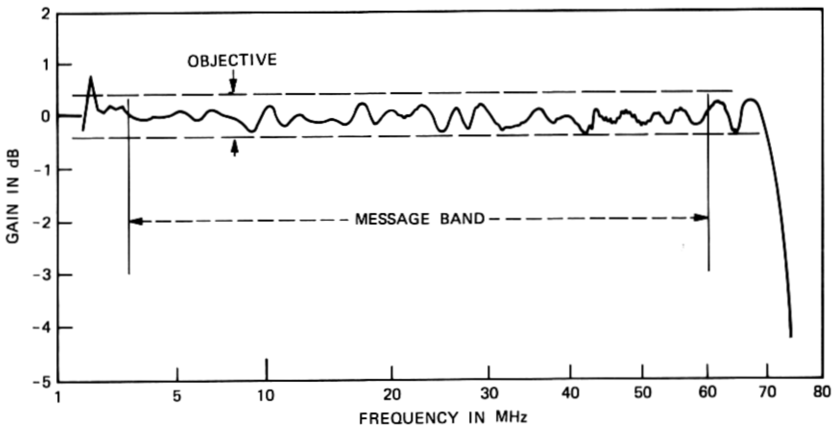

Fig. 35—Switching section equalized line response.

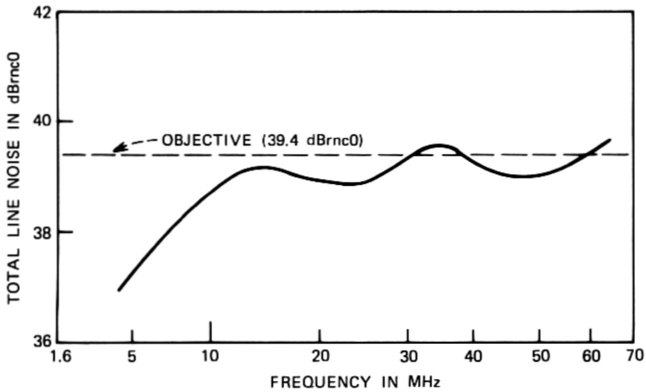

Fig. 36—Measured line noise (noise loading) extrapolated to 4000 miles (measured length: 750 miles).

reproducible response is a key to the success of the equalization strategy presented in this paper. The achievement of the objectives in all repeaters can be best described by Fig. 35, which shows the measured response of an equalized line. Also, noise measurements are compared with expected values and objectives in Fig. 36. These results indicate that the objectives of the repeatered line have been met.

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