Basic and Regulating Repeaters

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The line repeaters of the L-4 system represent a hierarchy of repeaters of increasing complexity: basic, regulating, and equalizing repeaters. Based on a building block philosophy, each of the more complex repeaters performs all of the functions of the less complex types plus additional functions of its own. The simplest, the basic repeater, is a plug-in unit with a shaped gain-frequency characteristic that compensates for the transmission loss of two miles of cable. The regulating repeater includes all of the components of the basic repeater plus two dynamic gain-regulating circuits and a deviation equalizer. Both repeaters include line building-out networks to compensate for departures from the nominal two-mile spacing. This paper develops the building block philosophy of the line repeaters and considers those aspects of the basic and regulating repeaters that contribute to load carrying capacity, reliability, noise figure, and modulation performance. It emphasizes the design features of the line repeaters that contribute to a fully-hardened system.

I. LINE REPEATERS

1.1 *Introduction*

The line repeaters of the L-4 system form a generic set of repeaters that are inserted at two-mile intervals along the cable route. All line repeaters perform the basic function of amplification; certain repeaters perform the functions of regulation and equalization. In the L-4 system, the primary functions of amplification, regulation, and equalization have been assigned in accordance with a building block philosophy to three types of repeaters. In order of increasing complexity, the (*i*) basic, (*ii*) regulating, and (*iii*) equalizing repeaters have been developed around a simple, fixed-gain repeater. Each of the more complex units performs all of the functions of the less complex types plus other special functions.

The fixed-gain basic repeater constitutes the main building block in the family of line repeaters. It consists, essentially, of two amplifiers—a preamplifier and a power amplifier—whose combined gain characteristics match the loss characteristics of the cable. Since the required two-mile spacing of the line repeaters cannot be applied rigorously because of geographic and other considerations, compensation is necessary. This is accomplished by providing line build-out networks, designed as fractional-mile artificial cables, to build out the cable to the nominal two-mile spacing.

This article considers only two of the three types of repeaters—the basic and regulating repeaters. The general considerations guiding the development of the basic repeater are reviewed and extended to include the regulating repeater.[1]

1.2 *Objectives and Design Considerations*

In the L-4 Coaxial System, 2,000 or more repeaters are required in a 4,000-mile circuit. The extent to which system objectives are met and maintained is largely dependent on the quality and reliability of the amplifiers used to compensate for cable attenuation. To guarantee the needed quality and reliability, stringent requirements are placed on the gain characteristics and on the quality of materials and components specified for construction. These, in turn, depend on current device and materials technology, on the techniques available for network design and fabrication, and on the current capabilities of high speed digital computer hardware and software.

Simplicity of design and the use of negative feedback are also important factors in maintaining close control of gain deviations. A design feature of the basic repeater, adapted from the L-3 system, is the absence of adjustable elements to control the gain characteristic.[2] By eliminating these gain-adjusting features, the design precludes the possibility of adjusting one element to compensate for the shortcomings of another and the possibility of introducing systematic errors by faulty or inaccurate adjustment. The adjustment of gain, required to accommodate for short-lengths of repeater spacing, is provided by the line build-out networks discussed above. These networks are plug-in units available in 0.1-mile increments from 0 to 1.0-mile lengths of cable.

In the basic repeater, the input and output impedances of the repeater are closely matched to the impedance of the cable in order to minimize the effects of echoes caused by line irregularities. Tolerable values of return loss at the repeater input and output terminals are typically 30 dB from 0.5 to 15 MHz, decreasing to 25 dB at 20 MHz. To minimize the effect of modulation products, the feedback circuits in the basic repeater are designed to be consistent with system modulation requirements by appropriate shaping across the transmission band.

In any repeater design, the ambient temperature of the repeater must be considered; the design and selection of components must be compatible with the permitted ambient temperature range. In the basic repeater, the design of the feedback loops of each amplifier is such as to control the repeater gain by providing a smooth, easily equalized gain shape, with adequate margins of stability, under conditions of aging and varying temperature.

Solid state repeaters operating at low voltage and current are more susceptible to damage by external disturbances than their vacuum tube counterparts. Therefore, new protection techniques and protection devices are required in the line repeaters of the L-4 system to protect against the following sources of high potential and transient currents

(*i*) Indirect lightning damage caused by induced longitudinals,

(*ii*) Short or open circuit transients caused by the interruption of the dc power source at voltages as large as 1800 volts,

(*iii*) Induced 60-Hz signals caused by proximity to, or faults on, high voltage lines,

(*iv*) Electromagnetic pulses of such a magnitude as to cause electrical breakdown of the coaxial cable or components within the repeater.

The unusual combinations of operating conditions such as high traffic density, large numbers of repeaters in tandem, relative inaccessibility of underground repeater locations, high voltage on the cable and repeaters, and the need for water-tight integrity of the apparatus cases, impose particular design emphasis on those considerations relating to reliability, personnel safety, and ease of maintenance. The selection of components and devices of known reliability, and conservative derating of the established capabilities of these items to survive electrical stresses, are mandatory.

Perhaps the greatest influence on the physical design of line repeaters is the requirement that the crucial routes be "hardened" sufficiently to survive the high overpressures of near misses by atomic weapons. Aboveground structures capable of meeting the hardness requirements specified for the L-4 system are inherently expensive.

The alternative of underground manholes is complicated by the requirement that watertight housings be provided, which places increased emphasis on small size and compact packaging. Transmission requirements and current semiconductor technology have resulted in fairly high power dissipation for the individual repeaters. The method of powering the repeaters imposes a high voltage to ground at repeaters which are close to power supply points. It is necessary, therefore, to provide high voltage insulation that has good thermal conductivity, and to maintain careful control over all other portions of the heat removal path.

Another complication imposed by the underground location is that manholes are potentially wet and dirty. These considerations led to a design requirement that all elements requiring maintenance be plug-in elements so that maintenance might be carried out in more favorable environments.

In L-4, the power is supplied over the center conductor of the coaxial cable in a series arrangement. The main disadvantage of series power operation is that removal of one element interrupts power to all others in the circuit. Interruption of the line power causes automatic shutdown of the high voltage converters at each end. Restoration of power, requiring coordination between the two ends of the circuit, is time consuming both in manpower and in circuit downtime. Removal of power also results in changes in equalization associated with the *A* equalizers in the line. As a result, re-equalization is required after power restoration. In order to avoid unreasonable loss of circuit time on each occasion that a plug-in amplifier is replaced, a power patching cord is provided for maintaining dc power continuity.

The regulating repeater has to meet all of the objectives stated earlier for the basic repeater and also has to compensate for changes in cable attenuation caused by temperature effects. The gain of the regulating repeater must be continuously variable, under the control of a regulator circuit, to compensate for these changes. The frequency response of the regulating networks is required to match the square-root-of-frequency loss characteristic of the cable over the transmitted band to within ± 0.05 dB. This accuracy of shape must be held over the entire regulation range.

II. BASIC REPEATER

2.1 *Basic Repeater Configuration*

Figure 1 is a block diagram of the basic repeater circuit configuration. This repeater consists of two separate negative-feedback am-

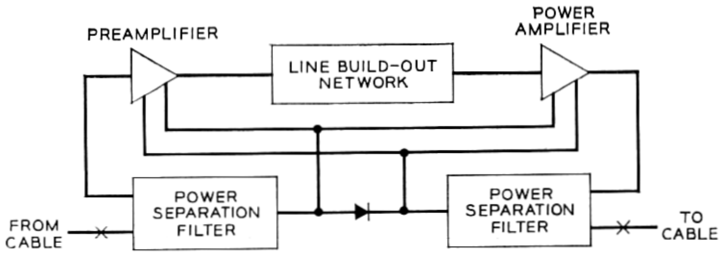

Fig. 1 — Simplified block diagram of basic repeater.

plifiers and a constant resistance line build-out network inserted between the two amplifiers. Both the preamplifier and power amplifier are connected to the coaxial line and isolated from earth ground by a power separation filter. An avalanche diode, reverse-biased by the dc line current, is used to provide a well-regulated voltage for the amplifiers. The required gain shaping of both amplifiers is accomplished in the feedback paths. Figure 2 shows the gain shape required of the repeater and the apportionment of the gain between the two amplifiers. This gain characteristic matches the loss of 2 miles of 0.375-inch coaxial cable at 55°F.

To realize the required repeater linearity and to provide margins against transistor aging and parameter dispersion, negative feedback is used. The most efficient feedback in terms of gain use, and perhaps the only practicable arrangement for this case, is major loop feed-

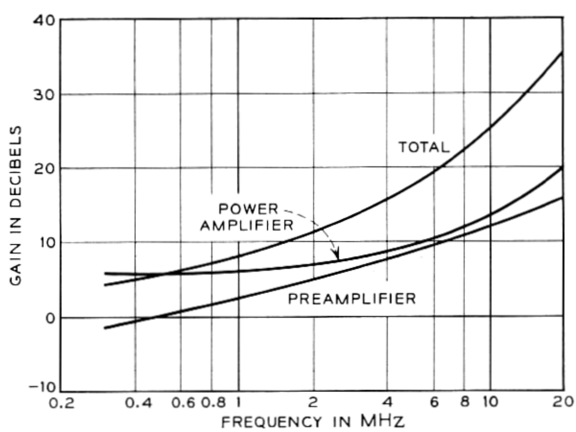

Fig. 2 — Contribution of preamplifier and power amplifier to total gain of basic repeater.

back. Since the two amplifiers use the same class of transistors, they tend to be limited to approximately the same maximum amounts of feedback at the top of the transmission band. High end loop gain considerations make it impractical to impose widely different gain burdens on the two amplifiers unless required by a lack of sufficient loop gain in one or the other. The preamplifier gain, therefore, is set in the region of 16 to 17 dB at 20 MHz, while the power amplifier gain is set between 18 and 20 dB at 20 MHz. The low frequency gain of the preamplifier is established by other factors which are dealt with in Section 2.2.4.

The two-amplifier repeater configuration offers several advantages: (*i*) the preamplifier can be tailored for low noise operation and the power amplifier for low distortion and high power output; (*ii*) excellent signal-to-noise performance is achieved because the gain is shaped within the feedback loop rather than by the introduction of lossy networks, either preceding or following the amplifiers; (*iii*) the feedback loops are relatively simple, and few loop gain shaping networks are required for the amplifiers.

Figure 3 is a photograph of the basic repeater. The frame of the repeater is a two-cavity, heavy-walled, H-shaped aluminum extrusion. Half of each cavity holds the components of the power separation filter; the other half holds an amplifier. The part of the frame containing the amplifiers is coated with 0.015-inch of epoxy to provide the necessary high voltage insulation.

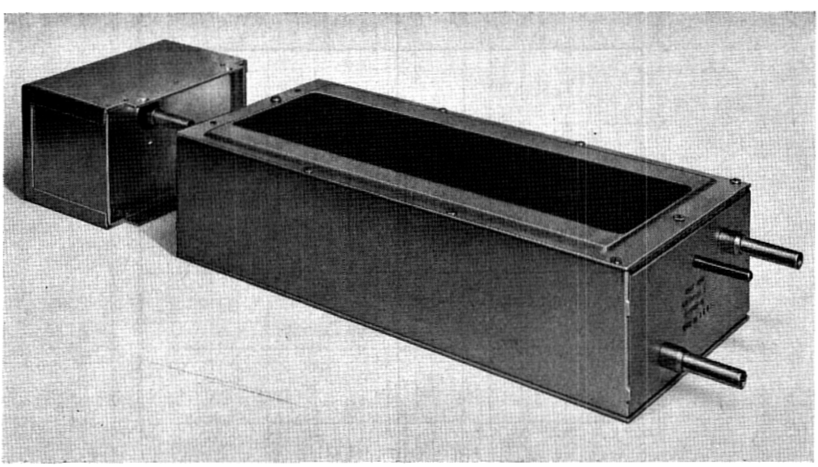

Fig. 3 — Basic repeater assembly with line build-out network (at left) detached.

The preamplifier and power amplifier are dimensionally identical. Figure 4 shows the preamplifier. For adaptation to other line and main station repeaters, minor alterations in package size and shape are required. Each amplifier has printed circuit construction. The printed wiring is etched in five-ounce copper bonded to the surface of a 0.062-inch epoxy-glass board. Printed wiring is desirable because of high operating frequencies and the resultant sensitivity to minor variations in component placement. The board assembly is fastened inside a heavy, die-cast aluminum box, open at the top and bottom. The sheet metal bottom cover is bonded to the epoxy coating on the repeater frame by a nylon-epoxy adhesive applied in sheet form to limit voids which are sources of corona impulse noise.[3]

Such construction ensures that most of the heat is conducted from the cover to the amplifier frame. It is then conducted to the repeater frame and eventually to the outside of the apparatus case. Figure 5 is a plot of the temperature gradients within the repeater. Notice that the gradients across the epoxy coatings (amplifier casting to repeater housing) are comparable with those across metal-to-metal interfaces.

In both the amplifier sections, thermally critical transistors are installed within heavy aluminum extrusions mounted to the printed wiring board and bolted to the aluminum covers. Only one of the two transistors in the preamplifier requires this form of construction. All three of the transistors in the power amplifier require similar treatment. In addition to providing a low impedance thermal path, the aluminum details serve to stiffen the board assembly by providing a rigid coupling to the top cover. Because of this structure the basic repeater can withstand 1.0 g vibration without exhibiting resonances between 5 and 500 Hz, and can tolerate shock levels as high as 250 g without damage.

2.2 *Preamplifier*

The preamplifier shown in Fig. 6 uses a hybrid feedback connection at the input and an emitter feedback connection at the output. The overall power gain of the amplifier is approximately 0.0 dB at 0.5 MHz and 16 dB at 20 MHz. The amplifier is highly linear and has a high frequency noise figure of 6 dB.

2.2.1 *Noise Figure*

In addition to the direct and obvious impact that noise figure has on overall system noise, the noise figure of the repeater determines,

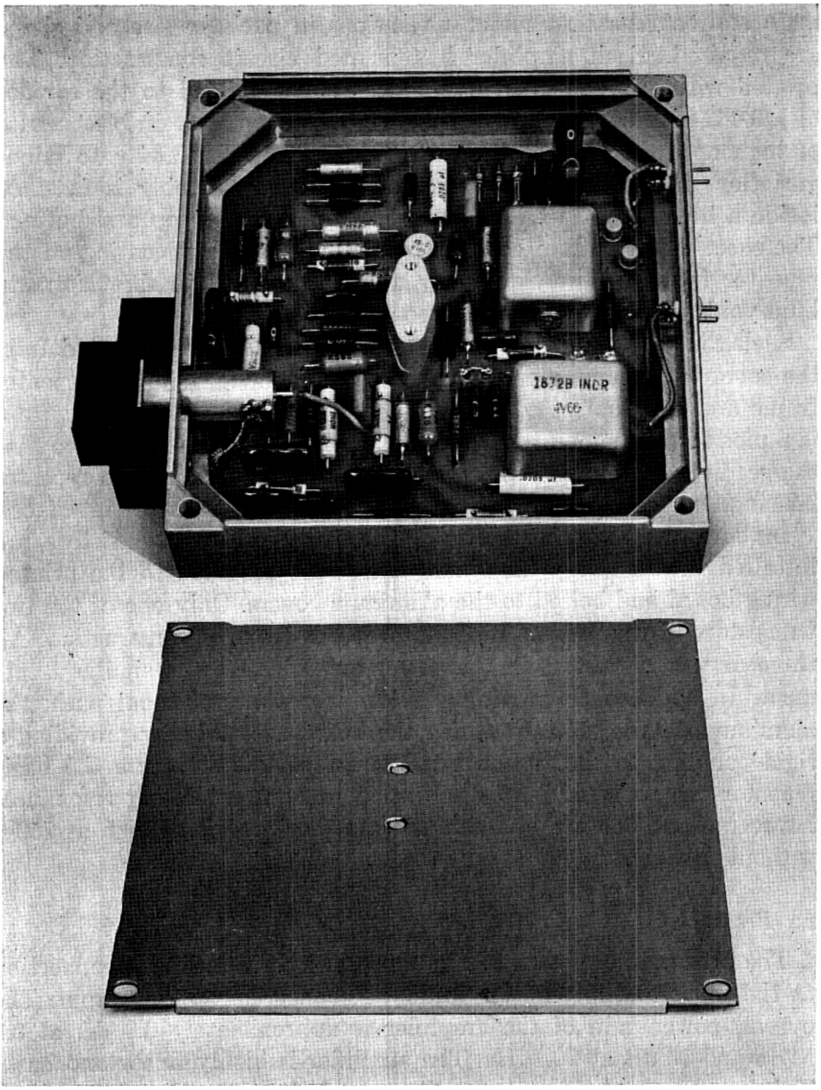

Fig. 4 — Top view of preamplifier with cover removed.

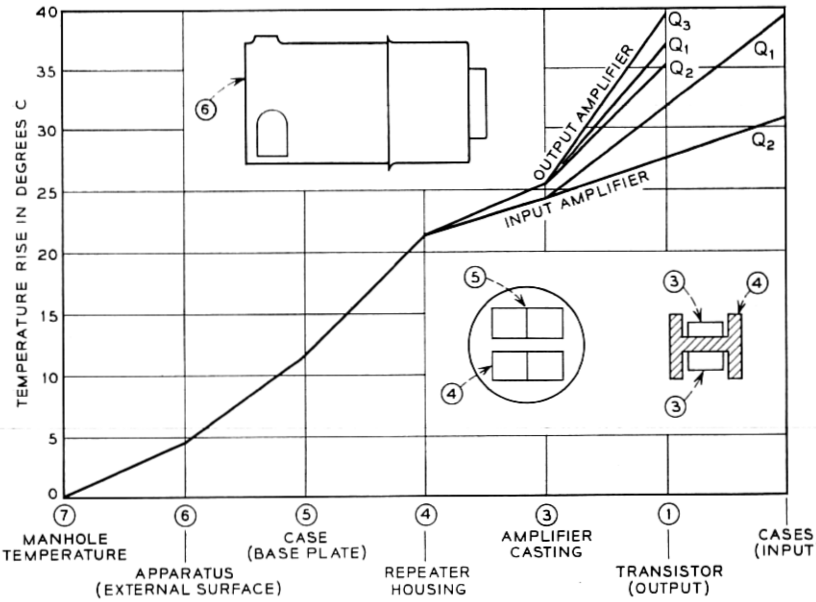

Fig. 5 — Temperature gradients for basic repeater installed in a manhole.

to a degree, the permitted load capacity. In a so-called overload limited system where modulation noise is no problem, the noise figure and system noise objective are the only factors of interest. The L-4 system, however, is modulation limited. As a result, transmission levels are fixed, not only by the repeater noise figure, but also by the nonlinear distortion indices.

The repeater noise figure is determined to a large extent by that of the preamplifier. Similarly, the preamplifier noise figure is determined primarily by the noise of the first transistor stage and the power loss of the input circuitry. The return loss and noise figure objectives suggest the use of a hybrid connection at the input if these objectives are to be simultaneously satisfied. Efficient use of device gain is mandatory if the several separate requirements are to be met. The first stage is a common emitter stage realizing both minimum noise figure and maximum device gain. Maximum device gain also serves to reduce the noise contribution of the second transistor stage and thereby minimizes the overall amplifier noise figure. The preamplifier input hybrid transformer has a nominal impedance ratio of 75:200 + 60. The hybrid loss associated with such a ratio is 1.1 dB. This loss,

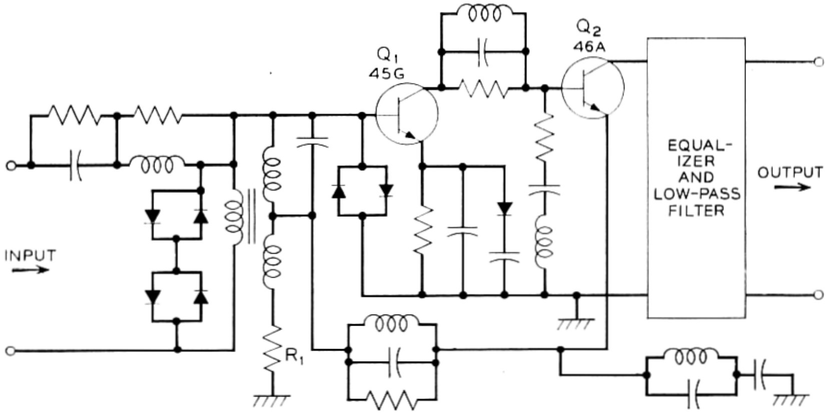

Fig. 6 — Schematic diagram of preamplifier (less bias circuitry).

plus the transformer dissipative loss of 0.4 dB, imposes a requirement of 4.5 dB noise figure for the input transistor if the amplifier noise figure is to be 6 dB.

2.2.2 *Beta Circuit Considerations*

Probably the most important factor in any amplifier design is the stabilization of the $\mu\beta$ loop transmission with satisfactory margins (about 30° phase margin; 10 dB gain margin). There are theoretical limits on the amount of feedback which can be applied to a particular structure when it is constrained to the use of a given set of devices and when selected gain and phase margins are to be realized.[4] In the preamplifier design, it is imperative that useful amounts of feedback, at least 10 dB, be maintained up to a minimum frequency of 30 MHz. The need for a hybrid connection at the preamplifier input places severe requirements on the hybrid transformer used in the feedback loop. In order that reasonably simple and practical circuit techniques be effective, the transformer design must be carried out with considerable care to ensure that no spurious resonances occur at frequencies up to 100 MHz. The measured $\mu\beta$ characteristic is shown in Fig. 7. The "points" superposed in the figure are the result of a nodal analysis, digital computation on the complete amplifier. The computed results and measurements agree within about 2 dB and 10° between 10 and 185 MHz. The discrepancies outside this range result from errors in device characterization, from errors in estimating values of parasitic elements, and from the slower velocity of propagation in printed wiring as compared to wire transmission.

2.2.3 *Mu-Beta Effect*

The closed loop gain of a single loop feedback amplifier can be expressed in the form: $1/\beta \cdot |\mu\beta/(1 - \mu\beta)|$; where β is the feedback circuit loss independent of device gain, and $\mu\beta$ is the loop gain. The magnitude of the term $\mu\beta/(1 - \mu\beta)$ that produces the so-called "$\mu\beta$ effect," is of interest primarily when the loop gain is small. Depending on the phase of the $\mu\beta$ loop transmission, the $\mu\beta$ effect may either increase or decrease the gain associated with "infinite feedback."

Case 1: For "$\mu\beta$" in Quadrant I or IV, there is a gain enhancement.

Case 2: For "$\mu\beta$" in Quadrant II or III, there is a gain reduction.

Case 2 applies at low and intermediate L-4 frequencies in the region below 10 MHz.

Case 1 applies over the upper portion of the L-4 band. In this range, increasing μ results in decreasing mu-beta effect, and the magnitude of the change depends on the particular $\mu\beta$ magnitude and phase involved in the computation of $\mu\beta/(1 - \mu\beta)$. Figure 8 shows the $\mu\beta$ effect in dB as a function of $\mu\beta$ phase for several values of $|\mu\beta|$.

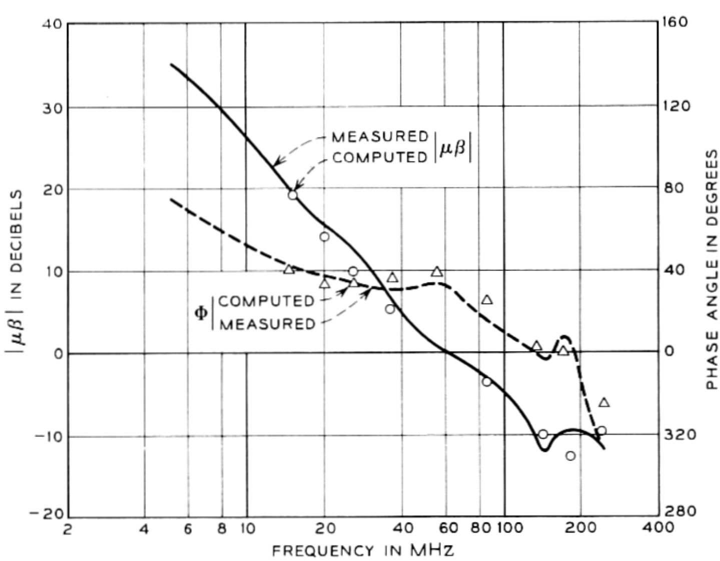

Fig. 7 — Typical measured and computed $\mu\beta$ or loop characteristic of preamplifier.

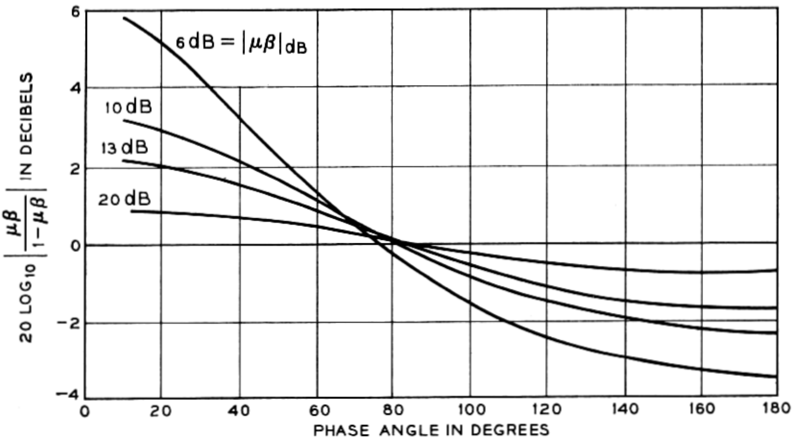

Fig. 8 — Calculated $\mu\beta$ effect vs $\mu\beta$ phase angle for various values of $|\mu\beta|$.

2.2.4 *Closed Loop Gain*

The closed loop gain of the preamplifier can be shown to be approximately

$$G'_v \cong \frac{n_3}{n_1} \frac{\alpha_2 \mid Z_L \mid}{\mid Z_{12F} \mid} \tag{1}$$

where Z_L = load impedance,
α_2 = common base short circuit current gain of the output stage,
Z_{12F} = open circuit transfer impedance of the feedback network.

If G_v is the voltage gain expressed in dB, and the input hybrid transformer turns are $n_1 = 7$, $n_2 = 3$, $n_3 = 10$, then $G_v = 3.1 + 20 \log (\alpha_2 Z_L / Z_{12F})$. To calculate the minimum amplifier gain, for a load impedance of 75 ohms and for a maximum value of Z_{12F} of 53.6 ohms

$$G_{v\,\text{MIN}} \cong 3.1 + 20 \log \frac{75}{53.6} = 6.3 \text{ dB}.$$

Because the amplifier design is not capable of achieving the desired minimum gain, the low frequency gain is reduced below the 6.3 dB computed above by placing an inductor in parallel with the output terminals. Additional low frequency equalization is also provided at the preamplifier output to achieve the overall repeater gain requirements. At 0.5 MHz, these networks reduce the gain of the preamplifier to about 0.0 dB, the 0.5 MHz goal.

Figure 9 shows the amplifier gain as measured between 0.5 and 20 MHz. The points plotted on the figure are the result of a computation using a nodal analysis computer program. The maximum difference between measured and computed values is 0.3 dB and occurs at 20 MHz.

It is evident from (1) that the gain of the amplifier is directly proportional to the α of the output transistor. This α-dependence tends to oppose the $\mu\beta$ effect variations whenever the $\mu\beta$ phase angle is in the first or fourth quadrants. At the low frequency end, where the phase of $\mu\beta$ is in the second or third quadrant, the α-dependence tends to supplement these variations; the $\mu\beta$ effect is generally small at the low frequency end where relatively high feedback is achieved. For relatively high feedback, the low frequency gain variations are proportional to the changes in the α of the output device. The β variations of 100 to 200 correspond to α variations of 0.990 to 0.995. The resulting variation in low end amplifier gain is ± 0.02 dB from some median gain. For β variations between 50 to 200, the corresponding gain variation is about ± 0.08 dB.

The effects of β_1 and β_2 on the amplifier gain is shown in Fig. 10. The gain variation caused by a changing β_1 results only in variations in the $\mu\beta$ effect, and thus the gain variation increases steadily above 7 MHz. Over the same region, the gain variation resulting from a changing β_2 is steadily decreasing to about 16 MHz. At this point,

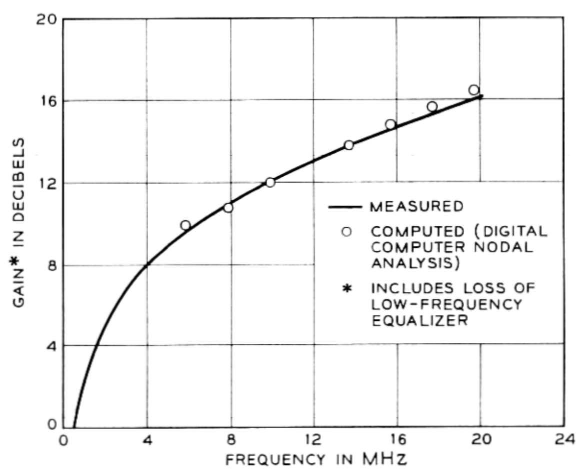

Fig. 9 — Measured and computed insertion gain of preamplifier.

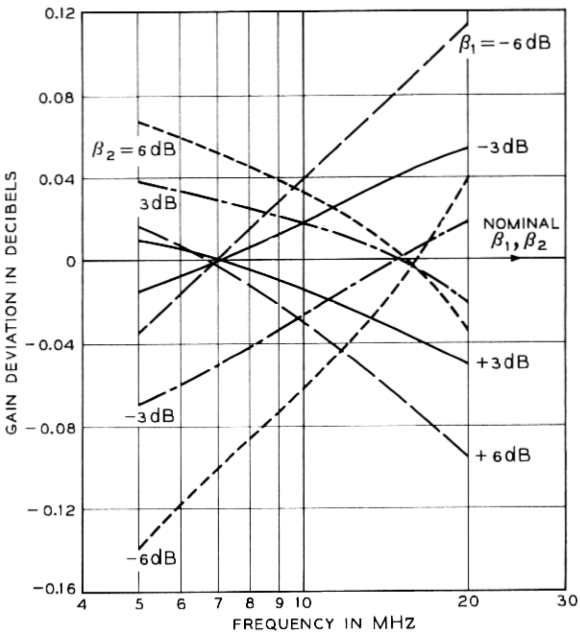

Fig. 10 — Calculated preamplifier gain deviations vs β of $Q1$ and $Q2$ transistors.

the change in α_2, as shown in (1), nearly cancels the change caused by the $\mu\beta$ effect.

2.3 *Power Amplifier*

The power amplifier, shown schematically in Fig. 11, is a three-stage, shunt-series, negative feedback amplifier. A common emitter—common emitter—common collector configuration was chosen as the optimum for intermodulation reasons. The amplifier provides a power gain that varies from about 5 dB at 0.5 MHz to about 20 dB at 20 MHz. The equivalent third order modulation coefficient at 17 MHz referred to 0 dBm, is about 105 dB. In addition to having excellent linearity, the amplifier can deliver up to +21 dBm into a 75-ohm load, with low distortion.

2.3.1 *Nonlinear Distortion Considerations*

The linearity of a feedback amplifier depends primarily on the inherent linearity of its transistors, the magnitude of the feedback, and the manner in which the feedback is applied. The transistor

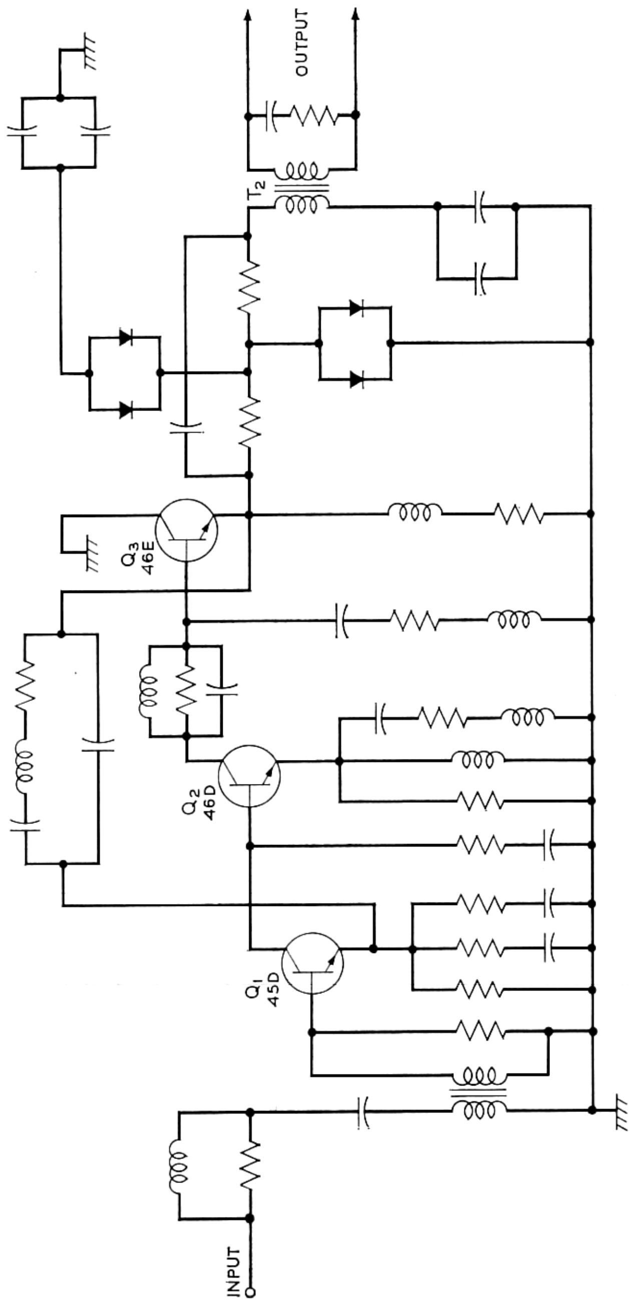

Fig. 11 — Schematic diagram of power amplifier (less bias circuitry).

linearity is, in turn, dependent on generator and load impedances. While the optimum load impedance for an output stage can frequently be provided, it is generally more difficult to optimize the driving impedance for such a stage without causing serious related impairments in gain. Maximum gain for a driver stage, in many instances, is not commensurate with optimum driving impedance for the output stage with respect to nonlinear distortion.

Comparisons of many transistor configurations and cascades, using a modulation test set developed for L-4, led to the use of the common collector output stage. The common collector stage has a third order modulation index about 5 to 10 dB better than the common emitter connection for the same load and generator impedances.

Because the power amplifier controls the modulation indices of the repeater, the transistor operating points are selected for maximum linearity consistent with reliable operation. This approach leads to operating points that may seem unrealistic from an efficiency standpoint; however, maximum linearity is the criterion.

Ideally, the nonlinear distortion should be controlled solely by the output stage. This would enable the circuit designer to predict the amplifier performance from a device measurement. Because of the low power gain of the common collector output stage, the second stage also contributes to the distortion. The third stage controls the third order modulation, while the second stage, because of its high impedance load, controls the second order modulation. The transistor operating points were determined experimentally to achieve the best balance between second and third order nonlinear distortion.

To predict the nonlinear behavior of the repeater over the L-4 frequency band, cross-modulation products are generated from three equi-amplitude, randomly-phased sine wave test signals. These test signals produce $f_1 \pm f_2 \pm f_3$, $2f_1 \pm f_2$ and $f_1 \pm f_2$ type cross products which can be related back to equivalent $f/3f$ and $f/2f$ modulation coefficients by the addition of appropriate constants. Laboratory measurements have shown that, when the feedback is moderate and frequency dependent, and when the transistor gain is also frequency dependent, conventional $f/3f$ and $f/2f$ measurements tend to give optimistic results.

In the conventional analysis of system modulation distortion, it is assumed that the output voltage of the open loop amplifier may be related to the input voltage by a power series:

$$e_0 = a_0 + a_1 e_1 + a_2 e_1^2 + a_3 e_1^3 + \cdots . \tag{2}$$

It is convenient to designate the power in dBm of the second and third harmonic components associated with a zero dBm fundamental measured at the output of the repeater or circuit under test. The modulation coefficients M_2 and M_3 may be expressed as

$$M_2 = f(a_1, a_2) \tag{3}$$

$$M_3 = f(a_1, a_3). \tag{4}$$

When the feedback loop of an amplifier is closed, the modulation coefficients M_{2A} and M_{3A} are defined by

$$M_{2A} = M_2 - F \tag{5}$$

$$M_{3A} = M_3 - F + K_F \tag{6}$$

where the subscript A in this instance refers to an amplifier. In (5) and (6) F is the feedback in dB, and K_F is a factor that accounts for the presence of closed loop third order distortion which would be present even if there were no open loop third order distortion whatever. The K_F factor exists if the second order products are fed back and modulated with the input fundamental signal, generating a third order interaction product. To simplify the analysis, K_F will be neglected in the following discussion.

If the power series coefficients are constant with frequency, and if the feedback is flat with frequency over the range of interest, then M_2 and M_3 are constants which may be designated as K_1 and K_2, respectively. When only the feedback varies with frequency, then,

$$M_{2A}(f) = K_1 - F(f) \tag{7}$$

$$M_{3A}(f) = K_2 - F(f). \tag{8}$$

Expressions (7) and (8) imply that all cross products of a particular kind which fall at a given frequency should be of equal power if the fundamentals are of equal power. Thus, a 1 MHz difference product formed from 16 MHz and 15 MHz fundamentals should be equal in power to a 1 MHz difference product formed from 3 MHz and 2 MHz fundamentals. This is, of course, a simplification. Reference 5 offers a more rigorous solution showing the dependence of the modulation product on feedback at frequencies other than the product frequencies. For example, in the case of an $f_1 + f_2 - f_3$ intermodulation product, the dependence extends to the feedback at each of the fundamentals f_1, f_2, and f_3 and at each pertinent interaction frequency, $f_1 + f_2$, $f_2 - f_3$, and $f_1 - f_3$, as well as the feedback at the frequency $f_1 + f_2 - f_3$. The nature of the process and not the mathematics is of interest here.

If the power series coefficients vary with frequency, it is then necessary to modify (7) and (8) as follows,

$$M_{2A}(f) = M_2(f) - F(f) \tag{9}$$

$$M_{3A}(f) = M_3(f) - F(f). \tag{10}$$

Equations (9) and (10) describe the modulation characteristics of the L-4 amplifiers with considerably greater accuracy than do (5) and (6). This can be attributed to the fact that the amount of feedback is not flat over the L-4 band, and that the frequency at which the common emitter current gain of the transistors begin to cut off ("Beta-cutoff") falls well below the top of the L-4 band. The Beta-cutoff for a 1 GHz transistor having an h_{fe} of 35 to 40 dB occurs at about 4 to 8 MHz; consequently, the transistors must be driven harder at high frequencies if a particular output power is to be achieved. This has the effect of increasing the distortion associated with input circuit variations and with the current transfer function. As a result, the "open loop" nonlinear distortion is severely frequency dependent.

Because of the cross product measurement techniques used, it becomes convenient to define "equivalent modulation indices," designated by M_{2E} and M_{3E}. The indices are equivalent because they are defined from measurements of cross modulation products rather than by a single frequency harmonic measurement. These indices permit the use of well established criteria for the calculation of system performance.[6]

The worst case modulation index can therefore be determined by using three high frequency tones that are close to the top of the transmission band where the system levels are the highest, and the feedback is a minimum, to generate an $f_1 + f_2 - f_3$ product that also falls close to the top of the band.

It is convenient and more representative of system performance to measure cross modulation products and calculate M_{2E} and M_{3E} by using the definitions:

$$M_2(f_1 \pm f_2) \text{ dB} \triangleq P(f_1 \pm f_2) - P_{f_1} - P_{f_2} \tag{11}$$

where $P(f_1 \pm f_2)$ = the second harmonic power at the output of the device or system in dBm at $(f_1 \pm f_2)$,

P_{f_1} = the fundamental power in dBm of f_1 at the output of the device or system,

P_{f_2} = the fundamental power in dBm of f_2 at the output of the device or system.

Similarly,

$$M_3(f_1 \pm f_2 \pm f_3) \text{ dB} \triangleq P(f_1 \pm f_2 \pm f_3) - P_{f_1} - P_{f_2} - P_{f_3} \,. \quad (12)$$

To calculate the equivalent indices, the techniques of Ref. 6 are used to arrive at

$$M_{2E}(f_1 \pm f_2) \text{ dB} \triangleq \begin{cases} M_2(f_1 \pm f_2) \text{ dB} - 6 \text{ dB} & \text{if} \quad f_1 \neq f_2 \\ M_2(f_1 \pm f_2) \text{ dB} & \text{if} \quad f_1 = f_2 \end{cases} \quad (13)$$

$$M_3(f_1 \pm f_2 \pm f_3) \text{ dB}$$
$$\triangleq \begin{cases} M_3(f_1 \pm f_2 \pm f_3) \text{ dB} - 15.6 \text{ dB} & \text{if} \quad f_1 \neq f_2 \neq f_3 \\ M_3(f_1 \pm f_2 \pm f_3) \text{ dB} - 9.6 \text{ dB} & \text{if} \quad f_1 = f_2 \neq f_3 \,. \\ M_3(f_1 \pm f_2 \pm f_3) \text{ dB} & \text{if} \quad f_1 = f_2 = f_3 \end{cases} \quad (14)$$

The measurements of both second and third order modulation indices are shown in Fig. 12. The plotted points shown on the figures are measured values at specific frequencies.

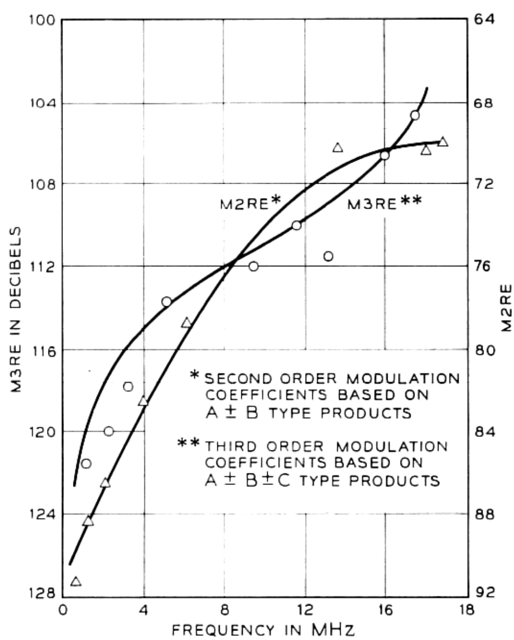

Fig. 12 — Second and third order modulation coefficients for power amplifier.

2.3.2 *Overload Performance*

Overload can be defined in many ways, depending upon the way the overload effect is observed when the amplifier is subjected to an increasing signal. Three common overload criteria for feedback amplifiers are (*i*) the "stonewall" effect, (*ii*) the change in modulation coefficient, and (*iii*) the change in gain effect. The "stonewall" effect applies to amplifiers with a very large amount of negative feedback. In such amplifiers, there is a point where a very small change in fundamental magnitude results in a very large change in harmonics. The modest amount of feedback at the top of the L-4 band (about 14 dB) eliminates the stonewall concept as a criterion. The change in gain concept, unfortunately, results in an overload point that is optimistic. The signal load that causes a significant change in the insertion gain of the repeater is approximately 6 to 7 dB beyond the point where the modulation noise begins to increase. Even the modulation coefficient effect must be examined critically and modified if it is to be used as a meaningful overload criterion for L-4.

Figure 13 shows how M_{3E} is affected by increasing the amplifier output signal power. It is apparent that the amplifier index degrades only gradually with increasing signal power, and that there is no

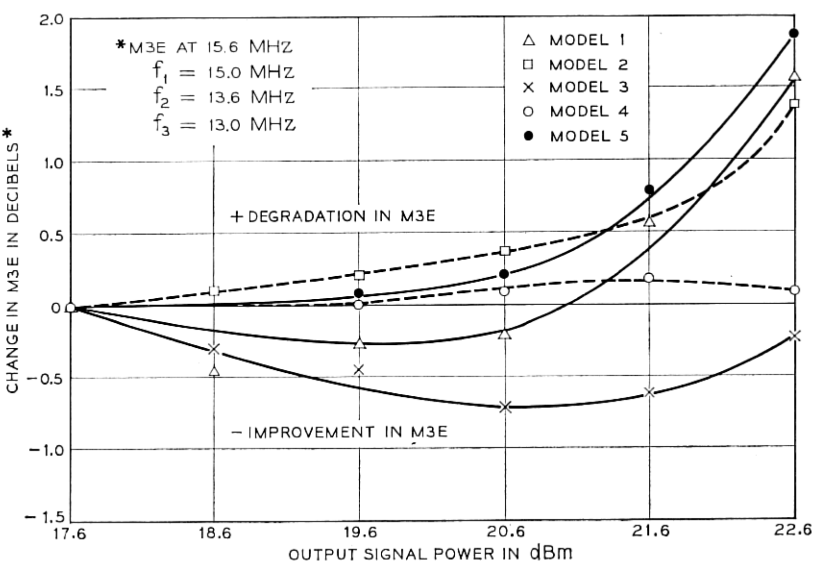

Fig. 13 — Change in third order modulation coefficient vs output signal power for power amplifier.

precise signal power in the interval measured where the performance becomes intolerable. This type of overload specification leads to a conservative design with an overload point as much as 6 dB lower than the other two criteria mentioned. In fact, as can be seen on Fig. 13, several models showed an improvement in M_{3E} with increasing output power. One model maintained an improvement in M_{3E} out to +23 dBm with respect to the index at +17 dBm. If it is assumed that the nearly constant index shown for signals below +20.5 dBm has been chosen for the L-4 repeater, then it is obvious that, if one amplifier or repeater fails to meet this objective by several dB, the effect on system noise will not be significant.

2.3.3 *Line Build-out Networks*

Line build-out networks are required in the L-4 system to build out the loss of cable sections shorter than the nominal two-mile spacing. Complications associated with route layout make it necessary to provide networks simulating the loss of cable in 0.1-mile increments from 0 to 1.0 mile. These line build-out networks are basically artificial lines designed to match the loss of fractional-mile lengths of cable.

Both artificial lines and cable equalizers may be designed by semi-graphical techniques involving the semi-infinite slope approximation.[7] Cable equalizers have been successfully realized by the tandem connection of a number of constant-resistance, bridged-T equalizers having parallel RC networks in the series arms.[8,9] The inverse problem of synthesizing cable simulators can be resolved by the use of similar constant-resistance networks having parallel RL networks in the series arms.

Although the use of the semi-infinite slope approximation is relatively simple, the visualization of the manner in which the gain and phase of the approximating function varies as the component poles and zeros are shifted about in the complex plane is more difficult. The problem is simplified considerably, however, in that the number of elements required to match a desired cable characteristic, with a given maximum ripple amplitude, can be determined before the actual design work is started. This simplification results from the observation that a minimum-phase loss ripple with an amplitude of one neper is accompanied by a phase ripple having an amplitude of approximately one radian, but which is 90° out of phase with the loss ripple.

If an infinite number of sections are assumed for the approximation, the determination of the loss ripple will require evaluation of an expression involving infinite products. The phase characteristic, on the other hand, is determined by an infinite series which converges rapidly. Since the two characteristics are related, the loss ripple is readily determined in terms of the phase ripple.

Performance data for the line build-out networks of the L-4 system are given in Fig. 14. These characteristics show the insertion loss of the 0.1 to 1.0-mile cable simulators. The number of bridged-T networks required for this simulation ranges from two sections for the 0.1-mile network to eight sections for the 1.0-mile network. An adjustable loss pad is also included in each of the fractional mile lengths (except the 0.1-mile section) to adjust the loss level to within ±0.05 dB of the required loss at 11.648 MHz. Although a total of 16 different bridged-T sections are required for the 0.1 to 0.5-mile designs, only a maximum of four sections are required for any one of these simulators. For the 0-mile line build-out network, a microstrip line is used to establish the required direct connection between the input and output jacks. A simple strap, at the frequencies involved, results in out-of-limit amplitude-frequency response.

The expression for insertion loss used in the design of the fractional-mile line build-out networks is given in (15).

$$\alpha(\mathrm{dB}) = 10 \log_{10} \frac{\left[1 + \left(\frac{f}{f_1}\right)^2\right] \cdot \left[1 + \left(\frac{f}{f_3}\right)^2\right] \cdots \left[1 + \left(\frac{f}{f_{2n-1}}\right)^2\right]}{\left[1 + \left(\frac{f}{f_2}\right)^2\right] \cdot \left[1 + \left(\frac{f}{f_4}\right)^2\right] \cdots \left[1 + \left(\frac{f}{f_{2n}}\right)^2\right]} \tag{15}$$

where f = frequency variable,

f_{2n-1} , f_{2n} = lower and upper break-point frequencies of the several infinite-slope approximations,

n = number of bridged-T sections (or number of semi-infinite slope approximations) required to meet the allowed tolerance on the matching function.

The line build-out networks are contained in aluminum sheet metal housings with epoxy-coated interior surfaces. They are firmly attached to the rear of the repeater housings by screws. Electrical connections are established by coaxial plugs which mate directly with the floating jacks at the output of the preamplifier and at the input to the power amplifier. Each of the bridged-T sections comprising the line build-out networks is mounted separately on individual printed wiring boards;

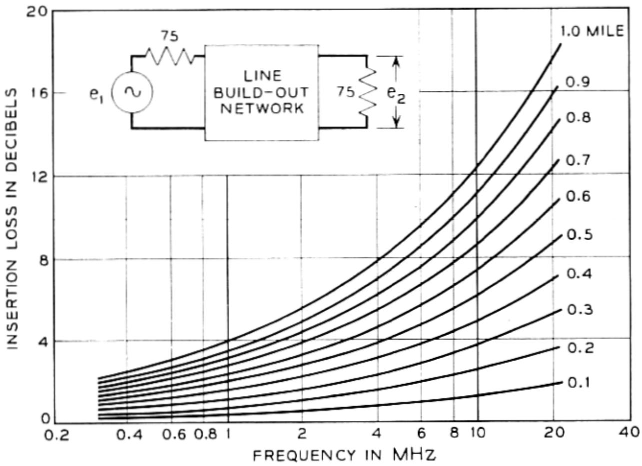

Fig. 14 — Insertion loss characteristics of line build-out networks (0.1 to 1.0 mile).

each is individually shielded. The units are adjusted separately and the appropriate designs selected to make up a particular fractional-mile cable simulator. The construction features of the line build-out networks are shown in Fig. 15.

2.3.4 *Transient Protection*

The line repeaters of the L-4 system must be able to withstand, without permanent damage, certain fault conditions which produce voltage spikes of several thousand volts and current transients of greater than 100 amperes for microsecond durations.

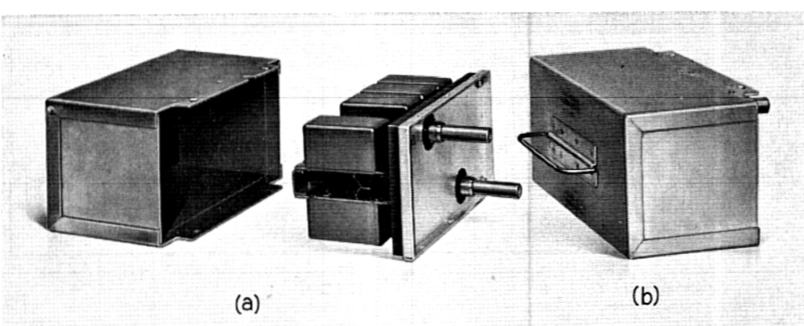

Fig. 15 — Mechanical construction of line build-out network: (a) network with cover removed; (b) completely assembled network.

The line amplifiers are protected by the silicon diodes in the circuits of Figs. 6 and 11. These circuit arrangements have been designed to satisfy the protection objectives for all types of known transients.

The limiting parameter that can cause device damage is the forward voltage drop across the base to emitter junction of the preamplifier first stage. To see the effectiveness of the protection circuits, examine Fig. 16. Figure 16(a) shows the voltage developed across the input of the line repeater caused by a momentary short circuit to ground of the 1,200-volt line at a repeater station two miles away. Peak-to-peak voltages greater than 200 volts are shown with frequency components between 100 and 500 kHz.

The effect of the primary protection diodes, shown on the line side

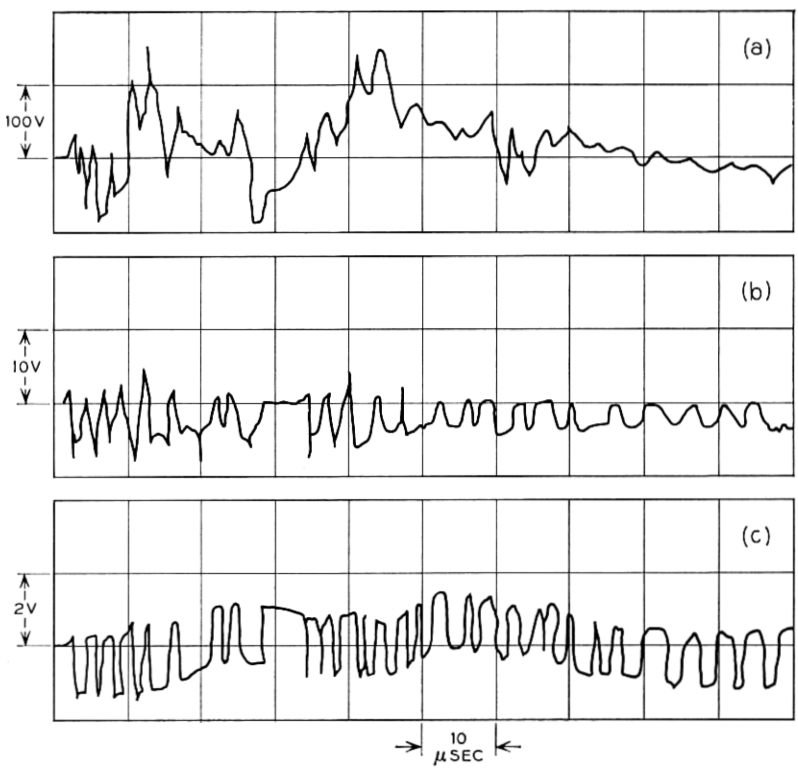

Fig. 16 — Transient voltages across (a) basic repeater input caused by a momentary short circuit applied to the high voltage line, (b) preamplifier primary protection diodes, (c) preamplifier secondary protection diodes.

of the preamplifier, is shown in Fig. 16(b). The voltage has been limited to less than 15 volts peak-to-peak.

The secondary diodes limit the peak-to-peak swing to about 2 volts peak-to-peak, as shown in Fig. 16(c). This is the maximum voltage applied to the first stage transistor.

The protection circuits described above also serve to protect against other transient conditions such as

(*i*) Lightning pulses with amplitudes exceeding 2,000 volts (it has been calculated that 600 to 800-volt protection should be adequate),

(*ii*) Induced 60-Hz longitudinal voltages of 850 volts at 3 amperes; test transients of this magnitude have been repeatedly impressed on an L-4 repeatered line with no adverse effect.

Extensive field work and testing in a radiation test chamber have demonstrated the adequacy of equipment shielding and the effectiveness of the protection circuits against electromagnetic pulses and resulting cable ionization.

III. REGULATING REPEATER

3.1 *Introduction*

The regulating repeater is the second in the hierarchy of repeaters of increasing complexity that constitute the line repeaters of the L-4 coaxial cable system.[10] The regulating repeaters are spaced at eight- to twelve-mile intervals along the system route and provide regulation and equalization to correct for deviations which can arise from the following situations

(*i*) geographic and other considerations which may alter the two-mile spacing of any type of repeater by as much as a mile,

(*ii*) predictable deviations accumulated from a number of basic repeaters in tandem which do not exactly match the cable loss over the frequency range of the system,

(*iii*) variations in cable loss resulting from changes in underground temperature along the system route. Typically, these temperatures can vary as much as $\pm 20°$F about a range of mean temperatures that extend from 40° to 75°F.

No provision is made in the basic repeater for gain adjustment other than that provided by the line build-out networks. In the regulating repeater, however, three sources of manual gain control are introduced: (*i*) flat gain adjustment of the regulating amplifier; (*ii*) level adjustment of an 11.648 MHz temperature control pilot;

and (*iii*) output voltage adjustment for initial alignment of an oscillator in the repeater preregulator control circuit. These gain controls are indicated in the block diagram of Fig. 17 which shows the additions to the basic repeater that are required to make up a regulating repeater.

To minimize system misalignment and to obtain a considerable signal-to-noise and overload advantage, both pre- and postregulation are used. For this reason, two line build-out networks are specified to permit independent adjustment of the post- and preregulating sections of the line. In the block diagram of Fig. 17, line build-out network 2 is shown adjacent to the power amplifier of the original basic repeater. In the actual circuit, however, this network appears between the hybrid and the temperature equalizer of the pre-equalizing section.

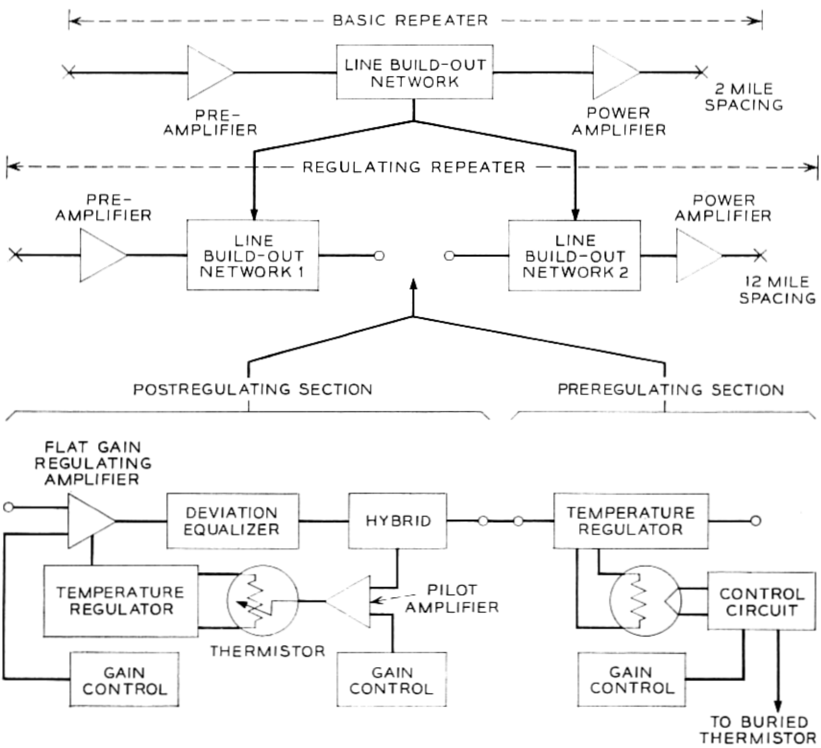

Fig. 17 — Simplified block diagram of regulating repeater showing build up from basic repeater.

3.2 *Regulation Philosophy*

The regulating repeater differs from the basic repeater only in its added equalizing and regulating characteristics. The functions of equalization and regulation have been introduced to correct for the two largest predictable sources of gain variation in the repeatered line (*i*) the variation of cable loss as a function of temperature, and (*ii*) the accumulated deviations from nominal gain-frequency shape of a number of fixed gain basic repeaters.

The largest single predictable source of gain variation in the L-4 system is that of change of cable loss with temperature. Locating the cable at an average depth of four feet underground reduces daily fluctuations in temperature to a point where they are practically nonexistent. Seasonal variations in some areas, however, can be as large as ±20°F about a nominal temperature. The magnitude of the loss change of the cable for an 18°F temperature variation about a nominal of 55°F is shown in Fig. 18.

The second largest deviation from nominal systems gain is the predictable variation accumulated by a regulating repeater and a number of basic repeaters in tandem along the line. These repeaters do not exactly match the cable loss over the entire frequency range of the system. Although the mismatches are small, the total accumulation for a maximum of six line repeaters in a 12-mile regulating section is appreciable. Constant-resistance deviation equalizers are available to correct for the gain deviations of three, four, five, or six basic repeaters.

3.3 *Block Diagram—Regulating Repeater*

The simplified block diagram of Fig. 17 was introduced to emphasize the development of the regulating repeater. A more complete diagram of the regulating repeater, including the power separation filters, is Fig. 19. The power separation filters are electrically identical to those of the basic repeater and are used to decouple dc power from the signal and to provide a high transmission loss from input to output at message frequencies. The high loss minimizes the possibility of unwanted feedback effects. The pre- and power amplifiers of the regulating repeater are identical to those of the basic repeater, but are packaged differently to conform with the overall design of the regulating repeater. The basic amplifiers essentially control the noise figure and modulation performance of the regulating repeater.

The postregulator is designed to introduce one half of the gain

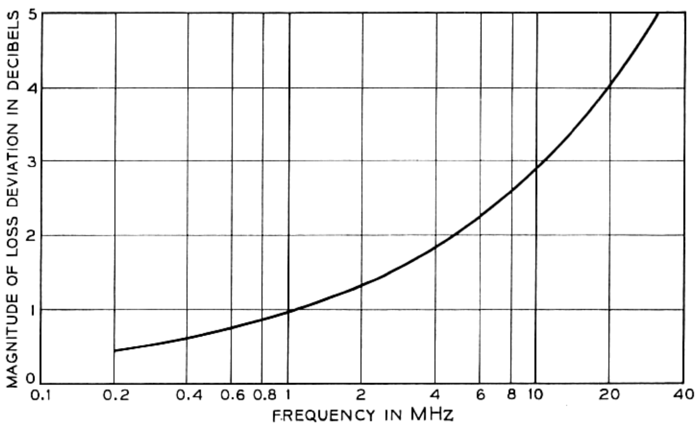

Fig. 18 — Magnitude of loss deviation resulting from an 18°F temperature change from nominal for 12 miles of coax 20.

correction required in a regulating section. This regulator is controlled by a continuous pilot signal at 11.648 MHz, located near the center of the transmission band between mastergroups 4 and 5. This tone is transmitted continuously over the cable at a level of −10 dBm0 and varies in response to changes in cable loss with temperature. The unequal ratio hybrid transformer following the deviation equalizer in Fig. 19 is used to direct the pilot tone to the closed loop of the thermistor-controlled regulator. After amplification and filtering, the tone is rectified, reamplified, and applied to a directly heated thermistor. Changes in the resistance of the thermistor present a varying resistance termination to a Bode-type regulator network. This network adjusts the gain in the control loop of the regulator by modifying the feedback, and thus corrects for temperature-associated changes in the characteristics of the cable. The deviation equalizer is introduced into the μ-path of the regulator to correct for the accumulated deviations of the line repeaters from nominal.

In the preregulating section of the repeater, a variable loss, Bode-type regulator network is used to correct for deviations in the cable characteristic resulting from temperature changes. A second thermistor, buried in the ground near the repeater manhole, is used to sense the cable temperature. The buried thermistor controls the output power of an oscillator in the control circuit shown in Fig. 19. Direct control of the associated Bode-network is established by still another

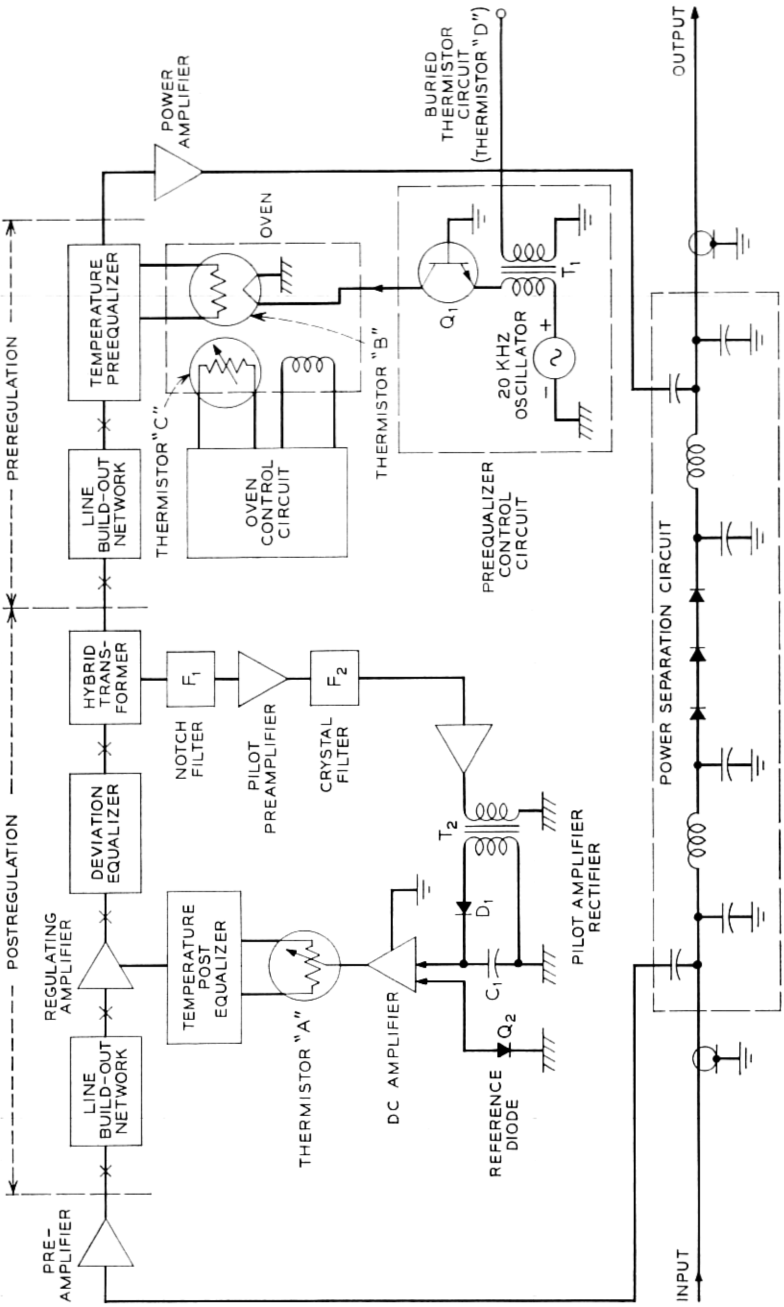

Fig. 19 — Complete block diagram of regulating repeater.

thermistor which responds to changes in the output level of the oscillator.

3.4 *Physical Design Considerations*

The physical design of the regulating repeater is closely related to that of the basic repeater. The detailed differences between the two are the result of disparity in circuit size and circuit complexity rather than differences in environmental requirements or changes in design philosophy. The added complexity, introduced to perform the additional functions, results in a package twice the volume of the basic repeater.

As in the basic repeater, the active circuits of the regulator may be as high as 1,800 volts above earth ground. To protect personnel and to insure against voltage breakdown, the same epoxy insulation as used in the basic repeater is applied to the interior of the repeater frame. This frame is divided into two sections, one for the regulating circuits, the other for the amplifiers. A top view of the regulating repeater, with covers removed, is shown in Fig. 20. The deviation

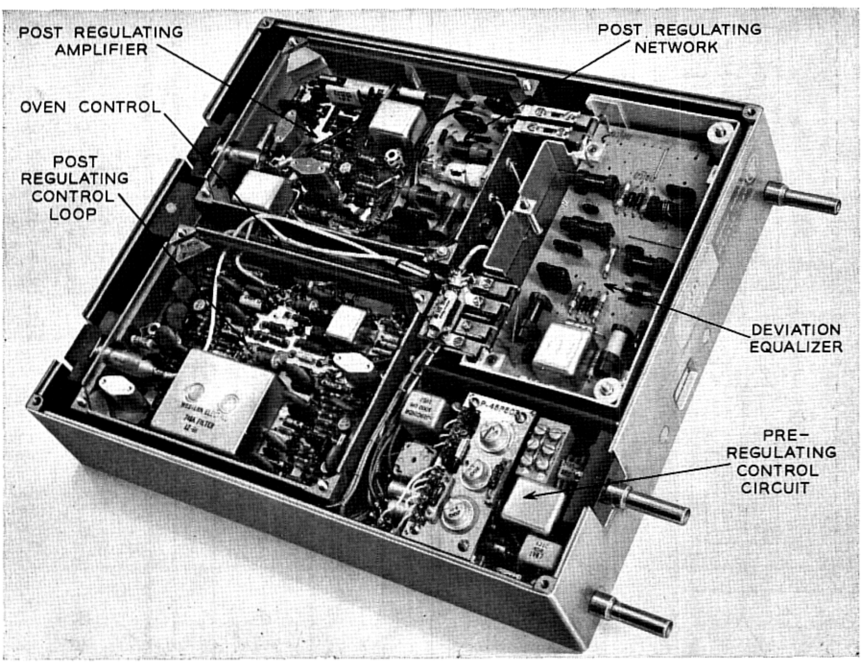

Fig. 20 — Top view of regulating repeater with cover removed.

equalizer, the preregulating control circuit, postregulating amplifier and network and the postregulating control circuits are identified. A bottom view of the repeater which includes the power separation filter and the pre- and power amplifiers is shown in Fig. 21.

In the regulating repeater, the pre- and postregulating circuits are packaged separately in heavy, die cast, aluminum frames, open at top and bottom. The bottom is closed with a sheet metal cover which is bonded to the repeater housing with an epoxy adhesive. Four mounting studs, welded to the cover, form mounting posts for assembly of the pre- and postregulating units in the repeater. A similar arrangement is provided for the deviation equalizer. The oven-control circuit, however, does not require a shielded enclosure. As a result, the printed wiring board assembly of this unit is mounted on four stand-off insulators. Similarly, a number of diodes, which must be insulated one from the other, are mounted on epoxy-clad steel brackets which are mounted to the repeater frame by nylon screws.

A completely assembled repeater, including the associated line build-out networks, is shown in Fig. 22. The repeater frame and cover, die cast from an aluminum alloy, have stepped, mating surfaces to insure accurate fit and to improve both the shielding and heat conducting properties of the assembly. The two external plug-in networks to the left in the photograph are the line build-out networks of the pre- and postregulating sections.

As discussed in Ref. 11, all of the line equipment of the L-4 system is designed for mounting in cylindrical, gas-tight cases. Each case contains internal framework to support the equipment and to provide the low impedance thermal paths required for efficient heat dissipation. The cylindrical cases accommodate four basic or two regulating repeaters.

The regulating repeater plugs directly into jacks mounted in the frame of the gas-tight case. Two of the three plugs shown on the front face of the regulating repeater of Fig. 22 permit the introduction of the repeater into the line. The third plug gives access to the buried ground-temperature sensing thermistor. Floating jacks at the rear of the repeater accept the plugs of the line build-out networks.

3.5 *Preregulator*

In discussing the features of the temperature-preregulator, it is instructive to re-examine the overall problem of cable equalization.

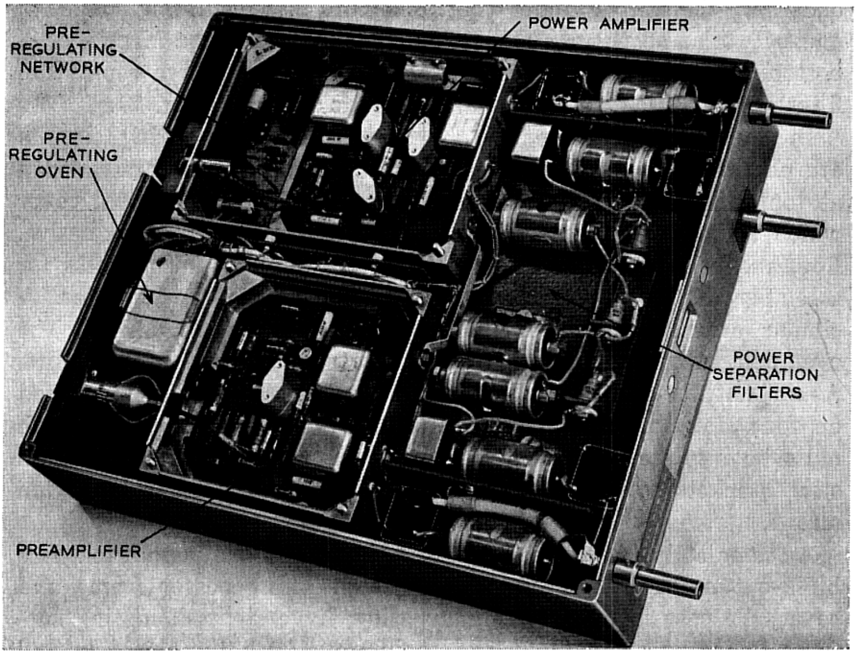

Fig. 21 — Bottom view of regulating repeater with cover removed.

An expression for cable loss is:

$$\text{cable loss (total)} = K_1 \cdot f(\omega) \pm g(T) \cdot f(\omega) \tag{16}$$

where

$K_1 \cdot f(\omega)$ = cable loss as a function of frequency,

$g(T) \cdot f(\omega)$ = cable loss as a function of both frequency and temperature.

The first term represents a loss that can be compensated for by fixed equalizers or shaped-feedback amplifiers. Correction of this term is assigned to the basic repeater. The second term indicates a loss dependence on both temperature and frequency, and correction is assigned to the regulating repeater. For regulation purposes, therefore, the cable loss to be corrected may be written

$$\text{cable loss (regulation)} = \pm g(T) \cdot f(\omega). \tag{17}$$

If one of a family of temperature dependent characteristics of the cable is selected as nominal—for example, that at 55°F—only deviations about the nominal need to be corrected. Assuming a passive network design, a fixed amount of flat loss is required to permit correction for deviations above and below the nominal. Since cable-temperature deviations are to be corrected equally between the pre- and postequalizers, the equation for the pre-equalizing loss may be written

$$\text{preregulating network loss} = K_2 \mp \frac{g(T)\cdot f(\omega)}{2} \tag{18}$$

where

K_2 is the flat loss of the network, and

$g(T)\cdot f(\omega)$ is equal and opposite in sign to the similar expression (17).

An insertion loss of the type required (18) can be realized by the basic series-type, Bode regulator network of Fig. 23.[12] This network, shown in detail in Fig. 24, consists of a four-terminal constant-resistance network and two associated resistors, R_1 and R_a · Resistor R_1 is termed the "symmetry" resistor and is of fixed magnitude. Resistor R_a is a variable control resistor whose magnitude is conveniently expressed in terms of the impedance level of the constant resistance network:

$$R_a = P(T)\cdot R_{01} \tag{19}$$

where $P(T)$ is a real variable which is a function of temperature.

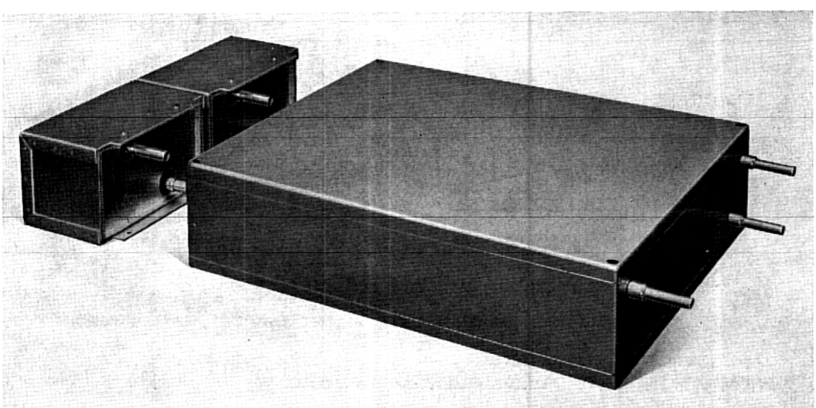

Fig. 22 — Regulating repeater assembly with two line build-out networks detached.

If the image impedances of the network are equal and of value R_{01} (for the case in which $R_a = R_{01}$), the insertion loss of the complete regulator will be independent of frequency. Further, if the insertion factor of the complete regulator is designated as ϵ^{θ} for the general termination $[R_a = P(T) \cdot R_{01}]$, and as ϵ^{α_0} for $R_a = R_{01}$, there exists a value of R_1 such that

$$\tanh \tfrac{1}{2}(\theta - \alpha_0) = \frac{K}{2} \rho \epsilon^{-2\phi} \tag{20}$$

where

$$\rho = \frac{1 - P}{1 + P}$$

ϕ is the transfer constant of the network,
and

$$K = 2 \tanh \frac{\alpha_0}{2}.$$

As pointed out by Lundry, Hakim, and by Ketchledge and Finch, simplifications in (20) may be made by using the series expansion of $\tanh^{-1}$ and by disregarding the high order terms.[13–15] If this is done, the loss of the regulating network may be expressed as

$$A = \alpha_0[1 + \rho \cdot R_e(\epsilon^{-2\phi})] \tag{21}$$

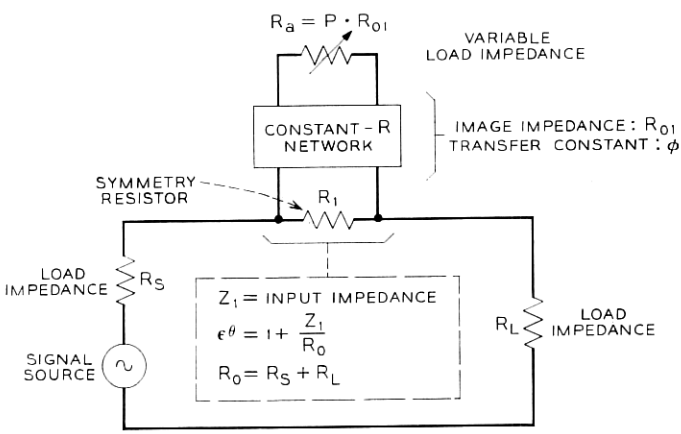

Fig. 23 — Series-type Bode regulator network.

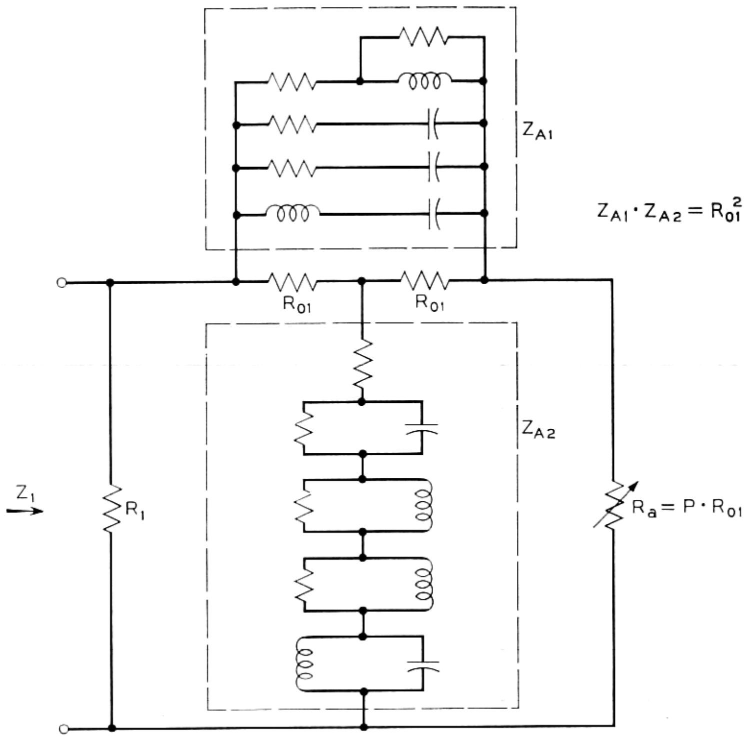

Fig. 24 — Preregulator network configuration.

where

A = insertion loss in dB,
α_0 = flat loss in dB,

$$\rho = \text{reflection coefficient} = \frac{P-1}{P+1},$$

Re = real part.

This equation and the two equations that follow are the design equations for the Bode network:

$$\alpha_0 = 20 \log_{10} \left[1 + \frac{R_1 \cdot R_{01}}{R_0(R_1 + R_{01})} \right] \qquad (22)$$
$$= \text{flat loss in dB,}$$

$$2\alpha_0 = 20 \log_{10} \left[1 + \frac{R_1}{R_0} \right] \tag{23}$$
$$= \text{max. loss in dB,}$$

and $R_0 = R_S + R_L$ = sum of source and load resistors for a series network.

Considering (21), the insertion loss is the sum of two terms. The first of these terms is a constant which establishes the reference level. The second is a function of the flat loss, of the reflection coefficient, and of the transfer constant of the shaping network. Since the reflection coefficient may be positive or negative, and since it varies with P, this second term adds or subtracts from the flat loss. It is therefore capable of producing a family of loss curves with mirror symmetry about the flat loss level. If (21) is normalized with respect to α_0, then

$$\frac{A}{\alpha_0} = A' = 1 + \rho R_e(\epsilon^{-2\phi}), \tag{24}$$

and the family of normalized curves has mirror symmetry about a flat loss of 1 dB. In this case, the minimum loss will be 0 dB and the maximum loss 2 dB.

The allowed tolerance of match for the preregulator is ± 0.05 dB over the entire frequency range of the system. The insertion loss of this network, operating between source and load impedance of 75 ohms, for three values of thermistor resistance, is shown in Fig. 25. When the magnitude of the thermistor resistance is R_{01}, a flat loss of 6 dB is established. The remaining two curves indicate the maximum and minimum correction available in the regulator for thermistor resistances of $3R_{01}$ and $R_{01}/3$. Intermediate values of cable correction are available between these limits for suitable changes in thermistor resistance.

3.6 *Postregulator*

3.6.1 *Deviation Equalizer*

The deviation equalizers included in the regulating loop of the postequalizing section of the regulating repeater are passive constant-resistance networks designed to correct for predictable deviations in gain accumulated by a number of fixed gain repeaters. As previously noted, equalizers have been made available to correct for gain deviations introduced by three, four, five, or six repeater sections making up a regulating section. (An n repeater regulating section consists of $n - 1$ basic repeaters plus one regulating repeater.)

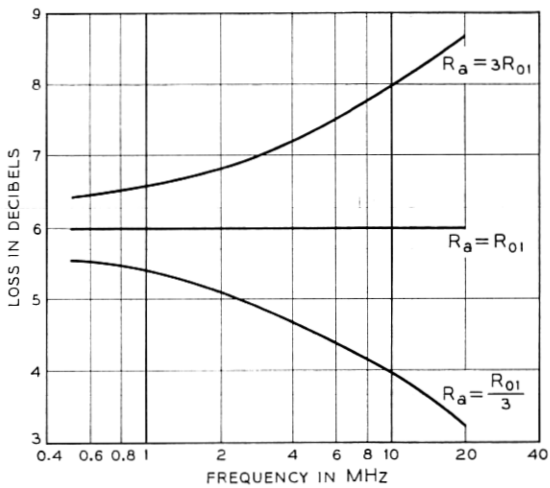

Fig. 25 — Insertion loss characteristic of preregulator network.

The deviation requirement for a five-repeater regulating section is shown in Fig. 26. A flat loss of 7 dB is permitted to meet the shaping and tolerance requirements. The tolerance on match is specified as ±0.05 dB over the entire frequency range of the L-4 system. The network configuration of the five-repeater deviation equalizer, designed to meet the foregoing requirements, is also shown in Fig. 26.

The mechanical design of the deviation equalizer is patterned after that of the pre- and postregulating sections of the regulating repeaters. A wrap-around frame, open at the top and bottom, is used to mount the printed wiring board assembly. The bottom cover plate is bonded to the insulated framework of the repeater with an epoxy adhesive. Three screws, forming mounting posts, are welded to the cover plate. The top cover is fastened to the equalizer frame by screws. The deviation equalizer with the top cover removed is shown in the photograph of the regulating repeater of (Fig. 20).

3.6.2 *Regulating Amplifier*

The regulating amplifier provides approximately 13.5 dB of gain from 0.5 to 20 MHz with a high degree of linearity and a good noise figure. As shown in Fig. 27, the design uses two common-emitter stages and a common-collector output stage. Figure 27 also shows the equation for amplifier gain.

Shunt feedback is used at the output and series feedback at the

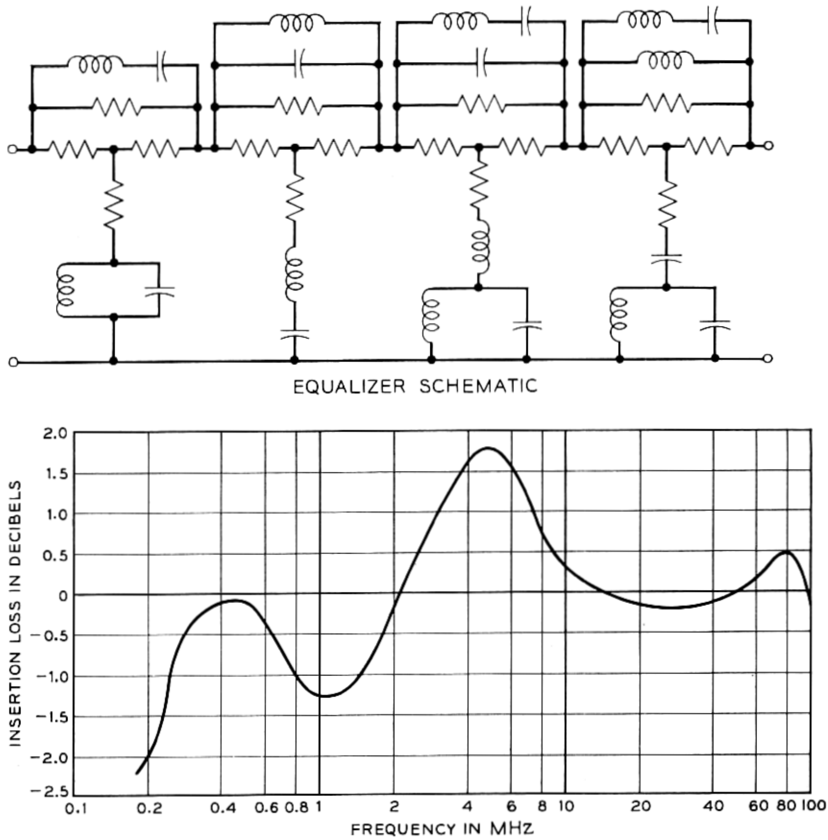

Fig. 26 — Deviation equalizer schematic diagram and requirements for a five-repeater regulating section.

input. As indicated in Fig. 27, the temperature postregulating network is placed in the shunt leg rather than the series leg of the feedback loop to insure stability under a wide range of two-terminal impedances. A terminated transformer is required at the input to provide a good 75-ohm termination for the preamplifier. This transformer adds 9 dB to the total voltage gain.

The gain adjustment feature, shown at the input to the first common emitter stage, is provided to compensate for flat loss deviations in the regulating repeater. The primary source of this flat loss variation is the variation in the flat loss of the six transformers used in the regulating repeater.

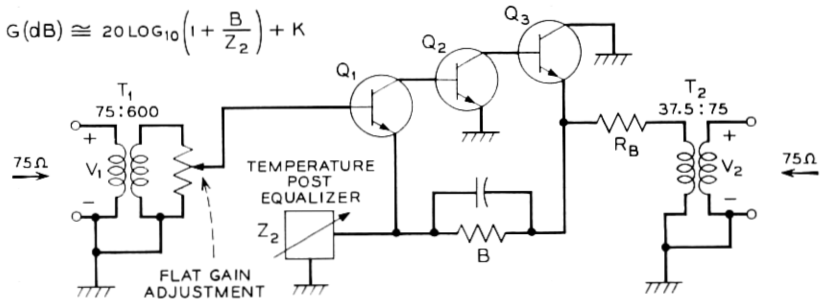

Fig. 27 — Simplified schematic diagram of regulating amplifier.

Design considerations relating to modulation products in the regulating repeater are identical to those for the basic repeater.

Other networks, not shown in Fig. 27, are required to obtain the open loop or $\mu\beta$ characteristics shown in Fig. 28. At least 40 dB of feedback is required at the low frequency end of the band and 15 dB at the top of the band to achieve the desired linearity and gain stability. The high frequency gain and phase margins are 7 dB and 30°, as shown in Fig. 28. Supplementary local feedback, not indicated, is also provided at each stage.

3.6.3 *Bode Regulating Network*

Section 3.5 points out that the postregulator must compensate for the remaining half of the cable-temperature loss deviation as given by (25):

$$\tfrac{1}{2}\text{ Cable Loss (Regulation)} = \tfrac{1}{2}\, g(T)\cdot f(\omega). \tag{25}$$

If the amplifier gain is such that it cancels this loss, then total equalization is achieved.

In Fig. 27, the gain of the regulating amplifier is expressed as

$$\text{gain (dB)} = 20 \log_{10}\left(1 + \frac{B}{Z_2}\right) + K. \tag{26}$$

The gain is therefore controlled by the ratio of B to Z_2. If B is fixed the gain is effectively a direct function of Z_2. Equation (26) may therefore be written

$$\text{gain (dB)} = 20 \log_{10}\left(1 + \frac{B}{Z_2}\right) + K = G(Z_2). \tag{27}$$

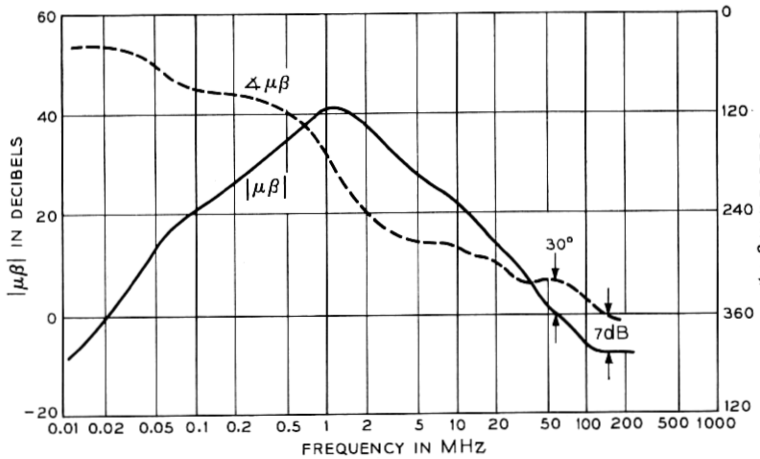

Fig. 28 — Typical $\mu\beta$, or loop characteristic, of regulating amplifier.

Complete equalization is achieved by equating (25) and (27)

$$G(Z_2) = \tfrac{1}{2}\, g(T)\cdot f(\omega). \tag{28}$$

In the development of the temperature preregulator, a series type Bode network was used directly in the signal path to precorrect for cable-temperature deviations. In the postregulator a shunt-type network, as shown in Fig. 29, is used in the feedback loop to correct deviations by shaping the gain of the regulating amplifier. Performance curves for the postregulating amplifier, with three different values of terminating thermistor resistance, are shown in Fig. 30. The accuracy of equalization is ± 0.05 dB over the frequency range of the system.

3.6.4 *Regulator Control Loop*

The regulator loop shown in Fig. 19 must sample, filter, and amplify the 11.648 MHz pilot tone without interfering with through transmission. A hybrid transformer is used to split the message band into two paths. One path connects to the preregulator section of the repeater, while the other path connects to a constant-resistance electrical notch filter.

The notch filter (F1) and its associated pilot preamplifier are inserted in the regulating loop between the hybrid and the crystal pick-off filter to minimize the so-called "nick" effect. The nick effect can be defined as an impairment in transmission caused by impedance irregularities introduced in the through transmission path by a bridging network.

The notch filter serves as the first step in pilot selectivity, prevents overload of the pilot preamplifier, and adds to the out-of-band suppression of the following pilot crystal pick-off filter. Figure 31 is a schematic diagram of the electrical filter and a graph of the insertion loss characteristic.

The pilot preamplifier, immediately following the notch filter, is a three stage negative feedback amplifier. It provides approximately 30 dB of gain and 25 dB of loop feedback at the pilot frequency.

Because of modulation and thermal noise in the system, a crystal filter (F2) is essential for pilot selection. This filter must have high out-of-band discrimination; it must also have good temperature stability to prevent excessive errors in regulation.

The pilot pick-off filter consists of two 180° crystal sections in a

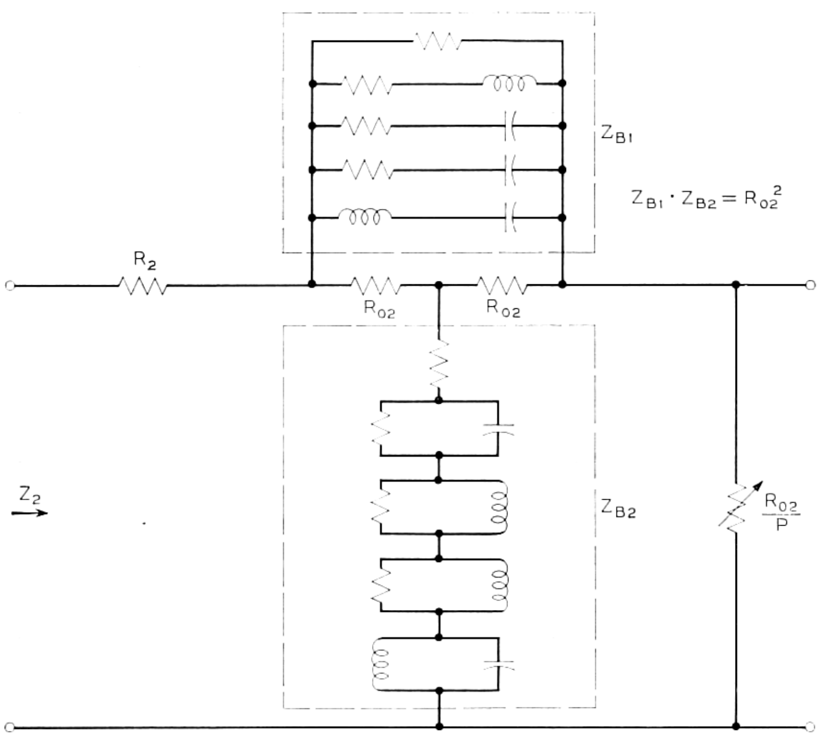

Fig. 29 — Schematic diagram of postregulator network.

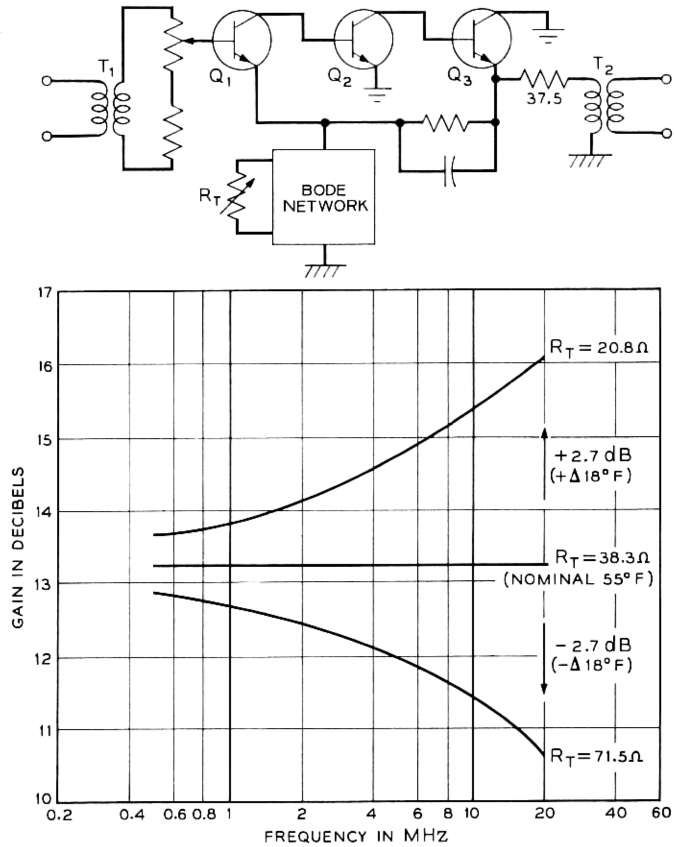

Fig. 30 — Performance characteristics of postregulator for a typical cable temperature swing.

hybrid configuration. The schematic and measured insertion loss characteristic of this filter are shown in Fig. 32. The inband characteristic, for several temperatures ranging from 30° to 140°F is shown in Fig. 33. The major shift in the characteristics is attributed to the change in the dc resistance of the hybrid transformers used to translate from a balanced lattice to the unbalanced hybrid configuration.

Packaging considerations, consistent with the miniaturization objectives of the transistorized system, governed the mechanical design of the filter. The filter, contained in its own package to permit maintenance of high out-of-band rejection, is mounted on the printed wiring board of the pilot amplifier. Fig. 34 shows the mechanical

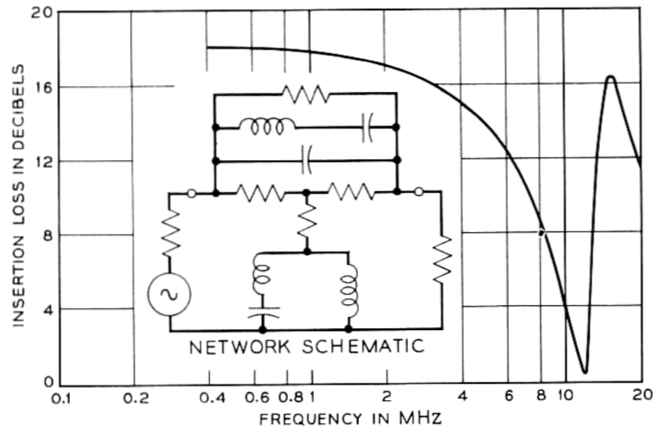

Fig. 31 — Insertion loss characteristic and schematic diagram of 11.648 MHz notch filter.

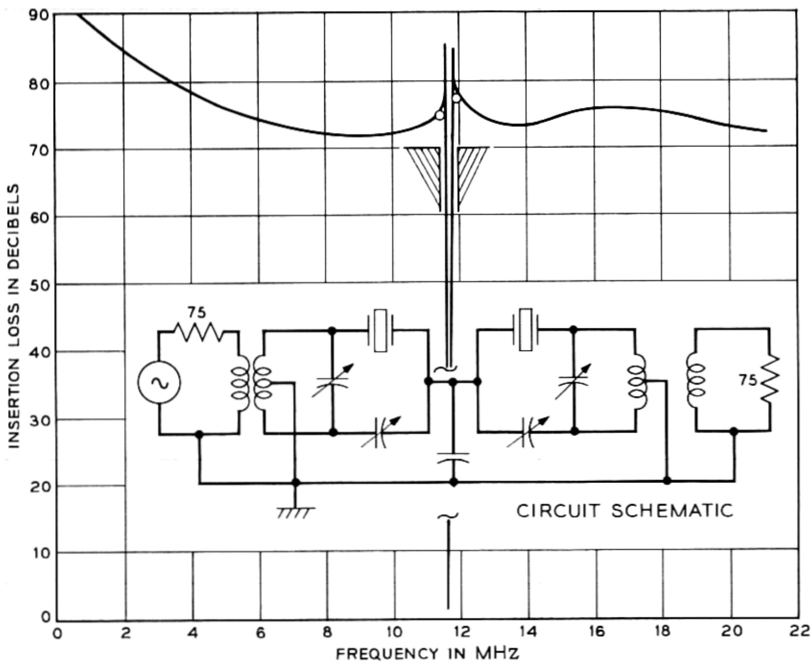

Fig. 32 — Insertion loss characteristic and schematic diagram of 11.648 MHz crystal pilot pick-off filter.

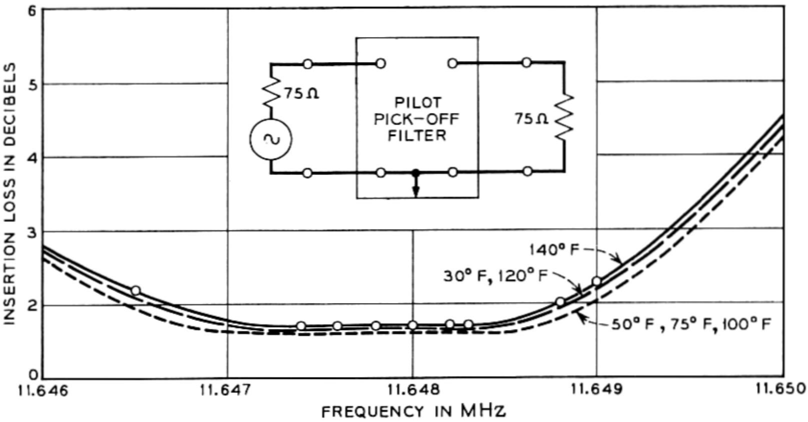

Fig. 33 — Inband temperature performance of pilot pick-off filter.

construction. The printed wiring board assembly is mounted to the cover by four standoff pins. The case has four recessed holes to allow adjustment of filter capacitors. After adjustment, the holes are sealed to make a hermetically sealed assembly.

Following the pilot pick-off filter, the pilot tone is further amplified and peak detected by the pilot amplifier-rectifier. A direct voltage, proportional to the amplitude of the pilot envelope, is obtained at the output of the peak detector circuit. A total gain of approximately 80 dB is required in the loop.

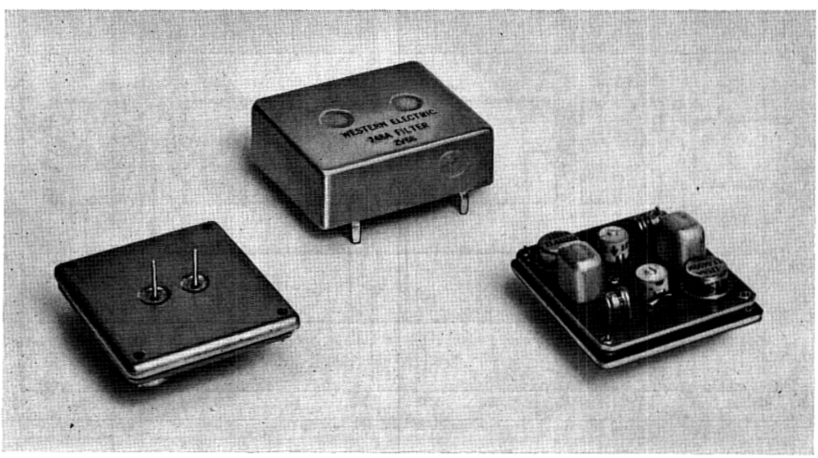

Fig. 34 — Mechanical construction of pilot pick-off filter.

A dc differential amplifier is used to compare the rectified pilot to a reference voltage obtained from a temperature-compensated diode. The difference voltage is then amplified. The output of the dc amplifier controls the resistance of a thermistor terminating the regulating network. The entire regulator loop is designed to keep the pilot output stable to ±0.1 dB over the operating ambient temperature range of the regulator.

A mathematical model of the regulator has been derived and is shown in Fig. 35. The following equations can be derived from Fig. 35.

Open loop transfer function

$$\frac{\Delta G}{\Delta e_0} = \frac{-K_3KK_1}{(ST_F + 1)^4(ST_T + 1)} = A\beta$$

Closed loop response

$$R = \frac{\Delta e_0}{\Delta e} = \frac{(ST_F + 1)^4(ST_T + 1)}{(ST_F + 1)^4(ST_T + 1) + K_3KK_1}$$

Where:

K = dc amplifier gain

$K_1 = \dfrac{\Delta e_B \text{ (volts)}}{\Delta e_0 \text{ (dB)}}$

T_T = time constant of thermistor

T_F = time constant of filter

$K_3 = \dfrac{\Delta \text{ regulating amplifier gain}}{\Delta I_T}$

This model was used for system performance evaluation and to obtain design requirements for components in the regulating loop.

The measured regulator envelope feedback characteristic is shown in Fig. 36. Envelope feedback of about 26 dB is obtained at low frequencies. The thermistor terminating the regulating network can be considered as a simple, low-pass filter with a cut-off frequency of 0.006 Hz because of its built-in time constant. It is important to consider that the extra phase shift introduced at higher frequencies

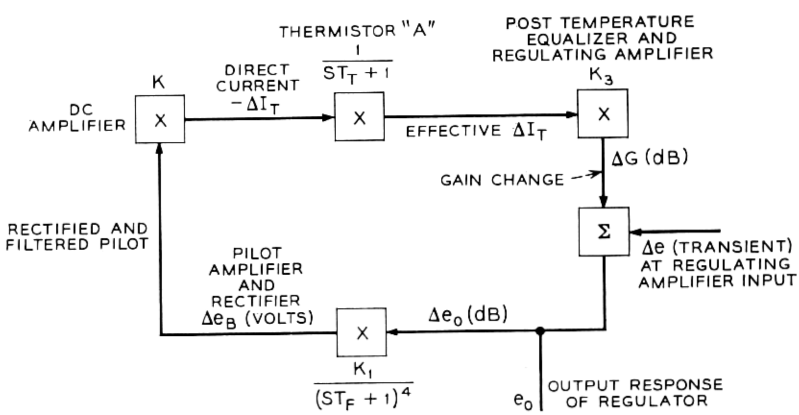

Fig. 35 — Mathematical model of the pilot-controlled postregulator.

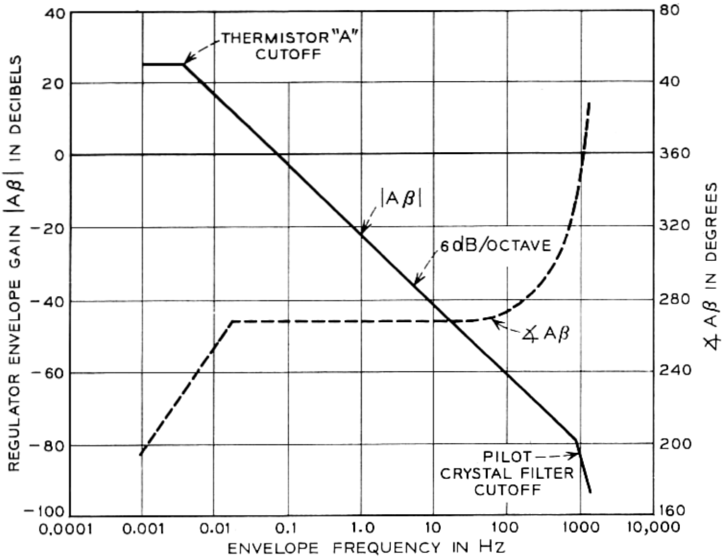

Fig. 36 — Measured postregulator envelope feedback characteristic.

by the pilot crystal filter can cause minute expansion in the pilot envelope. This is often referred to as gain enhancement.

Experience indicates that a system with envelope gain of less than 2 dB has a transient response that is nonoscillatory and well damped. The envelope gain enhancement of each regulator becomes important because as many as 500 can be used in tandem in a 4,000-mile system. The maximum gain enhancement occurs at approximately 400 Hz

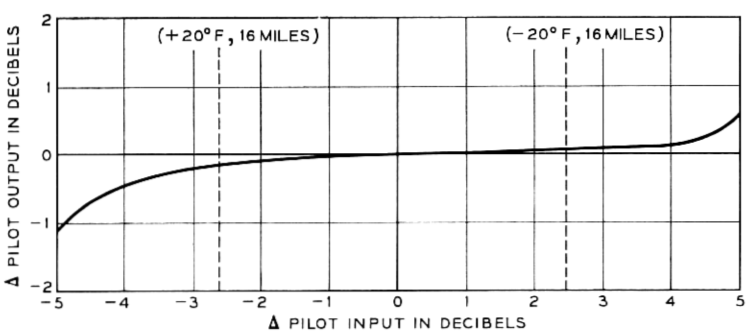

Fig. 37 — Regulating performance of pilot-controlled postregulator.

and is approximately 0.0013 dB per repeater or less than 0.7 dB for a 4,000-mile system.

The regulating performance of the pilot-controlled postregulator is shown in Fig. 37. As indicated there, the maximum expected change in pilot amplitude (about 2.5 dB) at the repeater input is reduced to less than 0.2 dB at the repeater output.

IV. SUMMARY AND CONCLUSION

The manufacture, installation, and subsequent in-service operation of the L-4 Coaxial System have confirmed the usefulness of the basic building block philosophy on which the development of the line repeaters of the L-4 system was based. Manufacturing operations have been considerably simplified by the concept of a hierarchy of repeaters in which the basic repeater is a fundamental building block. The use of line build-out networks to adjust for short lengths of repeater spacing has been successfully administered; the use of deviation equalizers to correct for predictable deviations in gain, as accumulated by a number of fixed repeaters, has been effective. Operationally, the noise, overload and regulating requirements specified for the line repeaters have been met and maintained. In summary, the line repeaters of the L-4 system represent a manufacturable product that has performed well within expectations.

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