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The L-4 Coaxial System

Foreword

This issue is devoted to the development of a new broadband coaxial transmission system which provides a more efficient and economic use of bandwidth on existing and new coaxial cables than earlier systems. The new system, called the L-4 Carrier System, provides 3600 two-way message channels on each pair of coaxials. Improved methods of equalization control and fault location have been provided and new multiplex equipment to form the signal spectrum for transmission over the medium has been developed.

The wide bandwidth and high performance have been achieved by using solid state circuits throughout. In addition to new transistors and diodes, other new components have been designed to provide extreme transmission precision and stability; the latest techniques in network design and synthesis have been brought to bear on the problems of network, equalizer, and filter developments.

Refined system and circuit design analyses have been used to obtain the maximum possible performance from all system components. As a result, the new system has about four decibels less noise than earlier coaxial systems. The L-4 system is now in commercial service and the field of application is expanding rapidly.

The articles in this issue discuss (*i*) the overall system and its design problems, (*ii*) the design of line repeaters and their broadband amplifiers, (*iii*) the techniques developed for equalizing the system, (*iv*) the methods of controlling equalizers and the new fault location system, (*v*) the development of new transistors and diodes, (*vi*) the terminal equipment required to interconnect L-4 with other systems and to provide interconnection between L-4 systems, (*vii*) the development of new power equipment for the coaxial line repeaters, and (*viii*) new outside plant developments including cable, manholes, and manhole apparatus.

Many people, too numerous to mention by name, throughout the Laboratories and other Bell System companies, have made significant contributions to the development of the L-4 system. Their efforts to bring the program to a successful conclusion and to complete the initial installation of the system between Miami, Florida, and Washington, D. C., on very short schedules must be acknowledged as having been outstanding and successful. Special mention must be made of the large and meaningful contributions of the late R. S. Graham (April 1915–October 1967) whose knowledge and efforts were so directly felt and appreciated in all aspects of the design of the L-4 System.

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The L-4 Coaxial System

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The L-4 Coaxial System is a new solid state broadband facility designed to transmit 3,600 long-haul message channels on standard 3/8-inch coaxial cable. This paper describes system design characteristics and operating features of the new system, discusses system power, protection switching, and fault location aspects, reviews the performance results from field trial and initial commercial service tests, and compares these results with predictions. This paper also summarizes the environmental factors and design concepts affecting the system's physical design.

I. INTRODUCTION

A new long-haul coaxial cable system, L-4, has been developed to transmit 3600 voice circuits on a pair of ⅜-inch diameter coaxial cables using frequency division multiplexing. A field trial was held in Ohio and completed successfully in early 1967. The system was turned up for commercial service in October 1967 on a route between Miami, Florida, and Washington, D. C.

The L-4 system is the latest in a family of long-haul coaxial systems. Each of the earlier systems was considered a large capacity system in its time.[1, 2] With the rising need for circuits and the advances in technology, each succeeding system has had increased capacity, as shown in Table I. Since system costs are dominated by the cost of the installed coaxial medium, the more efficient use of the line by the later systems effectively reduces the cost per circuit-mile. The increased capacity of the L-4 system, plus the relatively low cost of the solid state transmission equipment, has made it possible to reduce the cost of coaxial circuits by a factor of three compared with the cost of L-3 circuits.

In addition to its voice channel capacity, the principal features of the L-4 system are:

TABLE I—SYSTEM CAPACITIES

System designation	*L*-1	*L*-3	*L*-4
First commercial service	1941	1953	1967
Two-way voice channels per coaxial pair	600	1860	3600
Nominal repeater spacing (miles)	8	4	2
Typical route capacity per cable			
Number of coaxial units	8	20	20
Number of working pairs	3	9	9
Total two-way voice channels	1800	16,740	32,400

(*i*) All voice circuits satisfy the noise objective for 4,000 miles of transmission.

(*ii*) The system is completely solid state, thus minimizing maintenance and insuring maximum system reliability.

(*iii*) The nominal two-mile repeater spacing permits converting L-1 and L-3 systems to L-4 by adding repeater sites between those now used.

(*iv*) The L-4 frequency plan permits dropping any number of adjacent mastergroups without demodulating the other mastergroups.

(*v*) The number of continuous in-band pilots is minimized in anticipation of wideband analog and digital services.

(*vi*) All equalization adjustments are controlled from manned main stations, located as far as 300 miles apart, and can be made while the system is in service.

(*vii*) Fault location equipment is provided to locate a defective line repeater by measurements made from manned main stations.

II. FREQUENCY ALLOCATIONS

The line frequency spectrum for the L-4 system is shown in Fig. 1. The 600 voice channel mastergroup is a standard for long haul circuits in the Bell System. The six mastergroups are separated by 4 percent guard bands to facilitate dropping and blocking any number of adjacent mastergroups at a main station without demodulating the others. While this feature slightly increases the required bandwidth, it significantly improves the noise performance of the system by reducing the number of times that a through-mastergroup would otherwise be demodulated in a long circuit.

Even though it is not planned to transmit television on L-4, preliminary tests indicate that the same frequency allocation could be used for the television band as was used for L-3; the TV channel re-places mastergroups 2 and 3. This would permit the use of the television terminal developed for the L-3 system.[3] Specific operating problems will require further investigation if a need for television applications develops.

Two pilot frequency allocations are shown in Fig. 1. The only continuous in-band pilot is located at 11.648 MHz, in the guard band between mastergroups 4 and 5. This pilot is used in conjunction with regulating repeaters to eliminate, to a first order, the effect of cable temperature variations on the transmission characteristics of the system. A synchronizing pilot for the L-multiplex terminals is located at 512 kHz. This pilot, in conjunction with a band-edge regulator at main station repeaters, is also used to control the equalization of the low frequency end of the band.

A group of 12 command channels is located between 300 and 500 kHz. These channels are used for transmitting commands from manned main stations to the remote equalizers for system equalization. The band from 280 to 296 kHz is used for tones required by the protection switching system to coordinate line switches at both ends of a main station section. Finally, the frequency band from 18.50 to 18.56 MHz is used for 16 monitoring oscillator tones that are required for line repeater fault location.

III. SYSTEM DESCRIPTION—EQUALIZATION PLAN

3.1 *Basic Repeater*

Figure 2 is a block diagram of the L-4 system. The primary building block of the system is the basic repeater, nominally spaced at two miles. A fundamental principle used in designing the L-4 system was to keep the basic repeater as simple as possible. Consequently, this repeater has fixed gain, which is proportional to the square root of frequency, matching the loss at nominal temperature of 2 miles of ⅜-inch coaxial cable (see Fig. 3). The repeater consists of two fixed gain amplifiers and a power supply circuit. It has no gain adjustments but includes the option of a plug-in line buildout network which is selected when the repeater is installed. This network is used to compensate for differences in loss caused by repeater spacings less than two miles.

3.2 *Regulating Repeater*

The gain regulation necessary to compensate for changes in cable loss resulting from temperature variation is accomplished by inserting

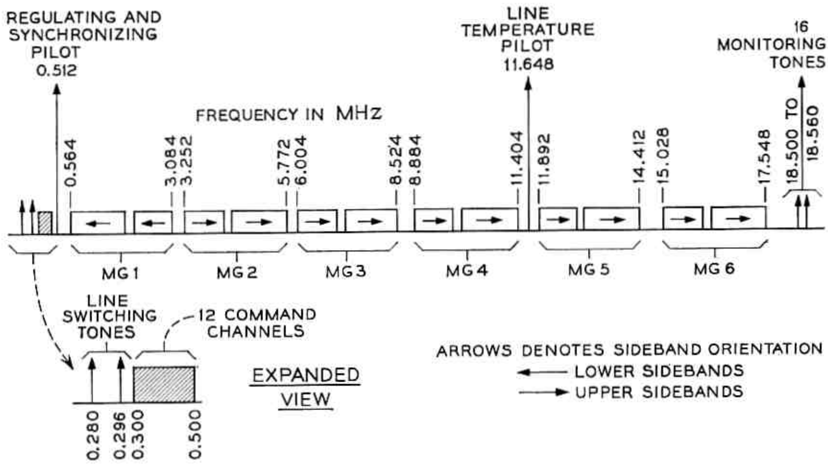

Fig. 1 — L-4 frequency allocations.

regulating repeaters in the L-4 line at maximum intervals of 12 miles. These repeaters include all of the components and features of basic repeaters and, in addition, include two dynamic gain adjusting circuits and a deviation equalizer.

One gain adjusting circuit, designated the postregulator, is controlled by the 11.648 MHz line pilot. This regulator varies the gain-frequency characteristic of the repeater to compensate for temperature associated changes in the loss characteristics of a regulating section—the cable and basic repeaters between two regulating repeaters. The second gain adjusting circuit, designated the preregulator, is controlled by a thermistor buried in the ground near the repeater in order to monitor ground temperature. This regulator preregulates by

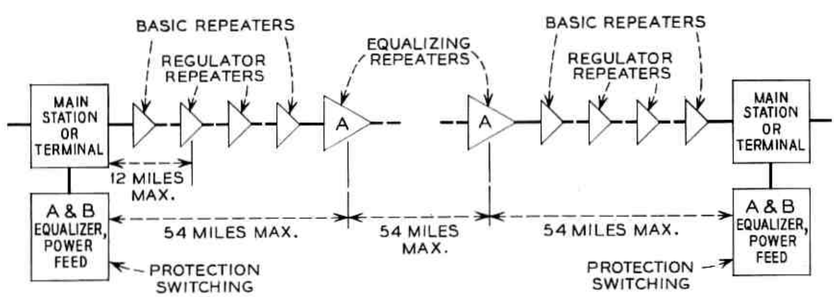

Fig. 2 — Typical L-4 repeatered line layout: maximum distance between main stations is 150 miles.

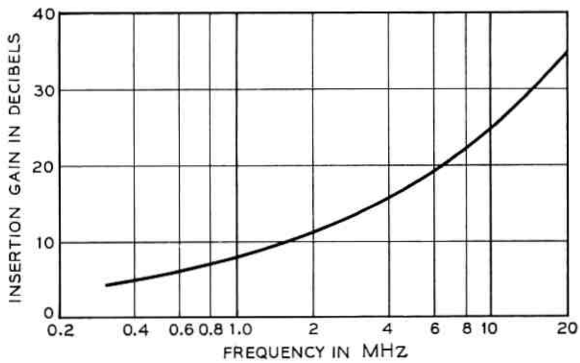

Fig. 3 — Repeater gain.

introducing approximately one-half of the gain-frequency correction required in a regulating section. The postregulator in the next regulating repeater introduces the other half of the necessary correction. By use of this technique of pre- and postregulation, it is possible to double the spacing between regulating repeaters without introducing excessive misalignment from cable temperature variations.

A deviation equalizer is included to compensate for transmission deviations caused by small but systematic deviations in the gain-frequency characteristics of basic and regulating repeaters from the loss characteristic of the coaxial cable. A family of deviation equalizers is required since the number of basic repeaters in a regulating section can vary between two and five. The deviation equalizer is a plug-in unit and is selected at the time the repeater is installed. Reference 4 gives a more complete description of the basic and regulating repeaters.

3.3 *Equalizing Repeater*

Equalizing repeaters, spaced up to 54 miles apart on the coaxial line, use all the components and features of regulating repeaters. In addition, these repeaters include six networks for adjusting the gain-frequency characteristic in order to compensate for random deviations introduced by the basic and regulating repeaters. The gain shape introduced by each network is effective over only a limited portion of the L-4 frequency spectrum. These networks, designated as an *A* equalizer, are adjusted remotely from manned main stations. These adjustments may be made while the system is carrying message service.

3.4 *Main Station Repeater*

Main station repeaters, spaced up to 150 miles apart, follow the building-block principles of the other repeaters. Each contains all the components and features of equalizing repeaters, ten adjustable equalizer networks, designated as a "B" equalizer, and a band edge dynamic gain adjustment. The sixteen adjustable A and B equalizer networks are remotely controlled from manned main stations. The band edge dynamic gain adjustment is controlled continuously by the 512 kHz out-of-band pilot. By means of the A and B equalizers and the band edge regulator in the main station repeaters and the A equalizers in the equalizing repeaters, it is possible to reduce the misalignment from all sources to about ± 0.5 dB in a main station section. Reference 5 has a full description of the equalizing and main station repeaters.

3.5 *Repeatered Line Power Equipment*

In addition to the repeater, main stations also include equipment for powering line repeaters and for protection switching. The power equipment, consisting of dc-to-dc converters, provides a constant direct current of 520 milliamperes. This current is fed over the center conductor of the coaxial unit together with the carrier signal. Depending on the distance between main stations, the dc voltage between the center conductor and ground can vary between 360 and 1800 volts. The primary power source for this equipment is the -24 volt office battery in the main station. The use of dc power, rather than the 60 Hz ac power used by the L-3 system, makes possible a substantial reduction in the complexity of the line repeaters as well as considerably simplifying the engineering of the system. Reference 6 has a detailed description of the power equipment.

3.6 *Protection Switching*

Protection switching equipment permits substituting a spare coaxial line for one of the working lines in either direction of transmission. A protection switch is automatically initiated by either the 512 kHz or the 11.648 MHz pilots going out of limits. It is necessary to make the switch between lines as rapidly as possible in the case of a total failure to minimize the effect of the transmission interruption. In order to coordinate the operation of the protection switching equipment at the transmitting and receiving ends of a main station section, signaling tones are transmitted in the 280 to 296 kHz band. Use of the high transmission speed of the carrier frequencies on the coaxials for transmission of these signals results in maximum speed of operation of the switching equipment.

For maintenance purposes, manual operation of the switching equipment is provided. In effect, manual switches are made by simulating a failure of the 11.648 MHz pilot.

3.7 *Manned Main Stations*

The equalization plan requires that at least every other main station be manned. Furthermore, a main station is usually manned if circuits are added or dropped at that location (terminal main station). A manned main station contains, in addition to the equipment previously described, a remote control system and terminal equipment as necessary for adding and dropping mastergroups. The remote control system provides a means for telephone craftsmen to measure the gain-frequency characteristics of the line, to adjust equalizer network settings at equalizing and main station repeaters, and to monitor fault locating signals associated with repeater stations along the line. A control center handles only receiving lines but may be used with up to 100 such lines: as many as ten cables, each with as many as ten receiving coaxials units.

Gain measurements are made at selected frequencies by means of single-frequency signals transmitted to the control center from the input or output of any selected equalizing or main station repeater. The tones are turned on or off and equalizer adjustments are made in response to coded commands generated at the control center and transmitted over the frequency band 300 to 500 kHz. Logic and memory circuits at the equalizing and main station repeaters receive, interpret, and respond to the commands from the control center and maintain the desired settings of the equalizer networks.

Commands from the control center can also cause fault locating signals to be transmitted from monitoring oscillators associated with any group of 16 line repeaters to the control center over the band 18.50 to 18.56 MHz. An individual repeater is identified by the unique frequency assigned to it in the group of fault-locating signal tones. Repeater failure or other malfunctioning can be detected by monitoring equipment at the control center. Reference 7 has a more complete description of the remote control system.

3.8 *Terminal Main Stations*

As previously mentioned, mastergroup multiplex equipment is provided in terminal main stations where circuits are to be added or dropped. The transmitting circuits accept six 600-channel mastergroup inputs from standard L-multiplex equipment in the band between 0.564 and 3.084 MHz, the so-called universal mastergroup band.[8] By suitable steps of modulation, each of the six mastergroups is placed in the appropriate part of the L-4 frequency spectrum. The resulting high-frequency line signal lies between 0.564 and 17.548 MHz, as shown in Fig. 1.

Receiving mastergroup multiplex circuits separate the mastergroup signals from one another, demodulating each one to the 0.564 to 3.084 MHz band. They are then delivered to receiving L-multiplex equipment for further processing as required. The terminal equipment is described in Ref. 9.

IV. SIGNAL-TO-NOISE ANALYSIS

4.1 *Transmission Objectives*

The L-4 4,000-mile zero level message noise objective, 40 dBrnC0, is 4 dB more severe than the 44 dBrnC0 L-3 system objective. The 40 dBrnC0 is allocated as 39.4 dBrnC0 to the repeatered line and 31.2 dBrnC0 to the terminals.

Signal-to-noise analysis and field evaluation of the L-4 system indicate that the random noise, linearity and peak power of the line repeaters, together with several noise reducing techniques described in Section 4.2, permit the L-4 system to meet the 4,000-mile 40 dBrnC0 noise objective.

4.2 *Noise Reducing Techniques*

Several system features affecting all of the repeaters are incorporated in the L-4 system to minimize noise. Because line repeater gain makes first circuit noise and third order intermodulation considerably greater at higher frequencies, signal shaping in the form of pre-emphasis of the high frequencies is used to ensure a fairly constant signal-to-noise ratio, thus suppressing noise more effectively at high frequencies.

A second noise reducing technique is negative feedback. The repeater feedback characteristic is designed to be at its maximum (about 37 dB) at low frequencies. Owing to feedback stability considerations, this leads to about 15 dB of feedback at 17.5 MHz, the top of the message band. This feedback frequency characteristic helps to compensate for noise introduced by intermodulation distortion. It also interacts with the pre-emphasis characteristic to produce a very nearly constant signal-to-noise ratio throughout the transmission band.

A third noise reducing technique is the use of pre- and postequalization. On every line section—between basic repeaters, between regulating repeaters, between equalizing repeaters, and between main stations—equalization is about equally distributed between the two ends of the section. Pre- and postequalization reduces the maximum misalignment resulting from accumulated deviations, thus eliminating the need for large changes in equalization at one repeater. This in turn reduces the possibility of overloading individual repeaters and minimizes noise penalties resulting from unavoidable transmission gain-frequency deviations.

A fourth technique for reducing noise is frogging. Frogging is the frequency transposition of parts of the signal transmission band at points along the repeatered line. Mastergroup frogging produces two major transmission advantages: (*i*) a break-up of the tendency of limiting third order modulation products of the "$A+B-C$" type to add in phase and (*ii*) a reduction in signal-to-noise penalties which result from strongly systematic misalignment along the repeatered line. In order to satisfy signal-to-noise and equalization objectives, all mastergroups must be frogged at intervals not greater than 800 miles. No through mastergroup should occupy the same frequency assignment in the L-4 spectrum for more than one frogging section in a 4,000-mile circuit.

4.3 *Thermal and Intermodulation Noise*

Predicted busy hour zero level noise in all channels is nearly the same, at a value of 39.4 dBrnC0 for the repeatered line, as shown in Fig. 4. At the top and middle of the band, the dominant modulation source is third order and is in optimum relationship 3 dB below thermal noise. At the low end of the band, the dominant source is second order, which is in the desired relationship when equal to thermal noise. Such optimum relationships, and the essentially flat overall total noise characteristic, are achieved primarily by means of pre-emphasis of the transmitted signal, which involves the control of signal magnitudes to match precisely the frequency characteristic of the total noise shape at a convenient point of reference.

4.4 *Signal Level and Repeater Spacing*

The dominant level shaping factor in a system such as L-4 is the insertion gain of the basic repeater, which is a fixed gain device de-

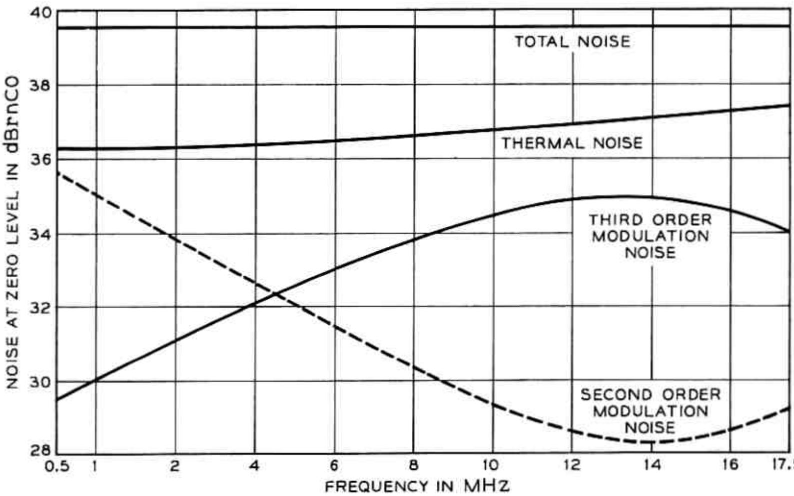

Fig. 4 — Zero level message channel noise for 4,000-mile L-4 high-frequency line.

signed to match the square root of frequency shape of the loss at nominal temperature of 2 miles of 3/8 inch coaxial cable. With reference to Fig. 3, gain varies from about 6 dB at the low end of the band to about 33 dB at the high end, resulting in a slope of 27 dB across the message spectrum. The second significant level shaping factor is the repeater noise figure-frequency characteristic; from the characteristic plotted in Ref. 4, the repeater noise figure ranges from 6 dB at the high end to 12 dB at the low end, which is in the direction of reducing the required signal shaping by 6 dB. Thus, the first approximation to the desired pre-emphasis is 21 dB. The final signal shaping is dependent on the details of the distribution and adjustment of modulation noise. A detailed signal-to-noise analysis indicates that 18 dB of pre-emphasis is optimum for the L-4 system. The amount of preemphasis is evident in Fig. 5, which shows the transmission level at the output of a line repeater.

Figure 6 shows a system block and level diagram, indicating design center transmission levels at points along the high frequency line.

4.5 *Misalignment—Margins*

In specifying transmission requirements for the line repeaters, margins are provided for signal to noise and overload penalties resulting from misalignment of signal levels as they are transmitted along the repeatered line. Provision is also incorporated for uncertainties which are present in the characterization of device and system parameters. Care is exercised in system design to minimize the required allowance for these effects in order to keep the requirements on the line repeaters as lenient as possible. Table II summarizes the allowances made for penalties and uncertainties in the L-4 system design. Measurements made in the Ohio field trial and on the Miami-Washington commercial system indicate that these allowances have been adequate.

V. ECHO CONSIDERATIONS

Analysis of the parameters that control echo performance indicates that for a 4,000-mile L-4 system, the long haul echo objectives will be met for the transmission of message, television, wideband data, and *Picturephone*® visual telephone service. These studies have assumed the use of the new and improved coaxial cable splice,[10] random cable reel length in accordance with repeatered line spacing rules, and a controlled sequence of cable manufacture and placement to optimize impedance match between coaxials at reel junctions. The analysis also takes account of the echo performance advantage, resulting from 800-mile mastergroup frogging, for message and data transmission.

The sum of echoes of all types in a 4,000-mile L-4 repeatered line is expected to be about 10 dB below the total allowable message system noise from all other sources. Repeatered line echo performance mar-

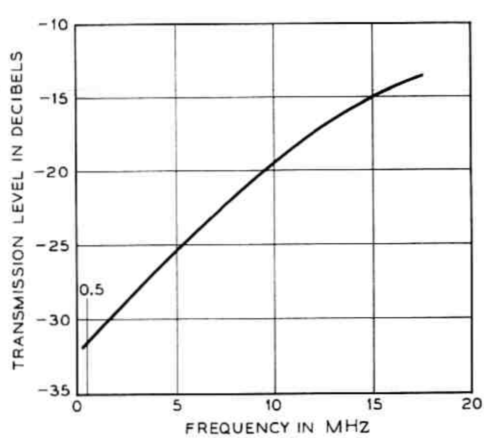

Fig. 5 — Transmission level at repeater output.

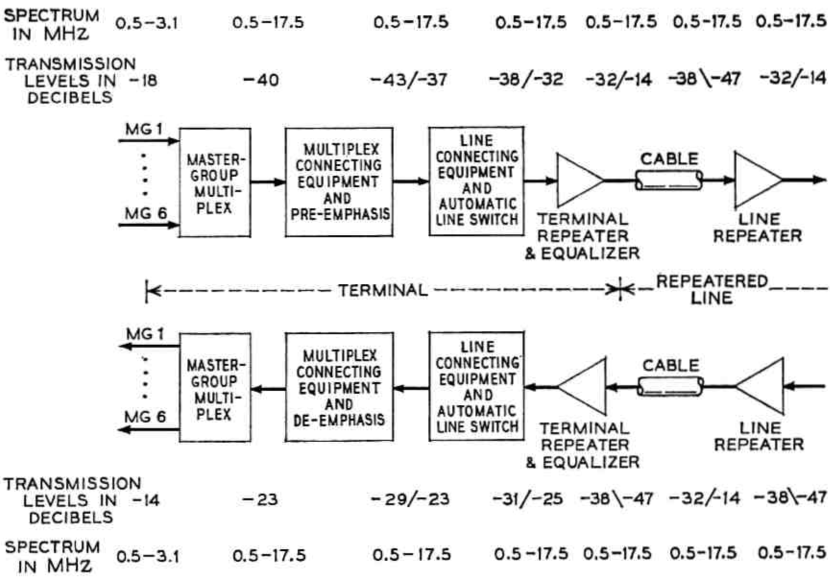

Fig. 6 — L-4 transmission levels.

gins of from 4 to 16 dB are predicted with respect to the appropriate 4,000-mile objectives for transmission of telephone, program, television, wide and narrow band data, and *Picturephone*® visual telephone service.

VI. FIELD RESULTS

This section summarizes the results of field transmission measurements made, initially on an L-4 field trial system in Ohio, and later on the first commercially installed L-4 system between Miami, Florida, and Washington, D. C. The overall results of these tests indicate that system noise, crosstalk, and transient characteristics are entirely satisfactory and in good agreement with theoretically predicted per-

TABLE II — ALLOWANCES FOR SIGNAL-TO-NOISE AND OVERLOAD PENALTIES BECAUSE OF MISALIGNMENT AND UNCERTAINTIES

Type of noise	Allowance (dB)
Thermal noise	2.5
Second order modulation	2.5
Third order modulation	5.0

formance. Line equalization capability meets the objective for a 150-mile main station link. Experience with the repeater monitoring oscillators shows that the method provides good capability for fault location and incipient trouble detection in the line repeaters.

6.1 *Signal-to-Noise Ratio Performance—Noise Loaded "V" Curves*

A noise-loaded "V-curve" performance extrapolated to 4,000 miles for a representative high end channel is shown in Fig. 7. Comparison of the measured and theoretical plots of Fig. 7 indicates that measured thermal and modulation noise components are in good agreement with theory, and that system transmission levels are at virtually optimum values for a fully loaded system.

6.2 *Impulse Noise*

Impulse noise performance margins are adequate under conditions dominated by extremely short duration impulses deriving from corona-like discharges in cable, connectors or repeater equipment, or dominated by impulse peaks generated by fully loaded systems. The corona-

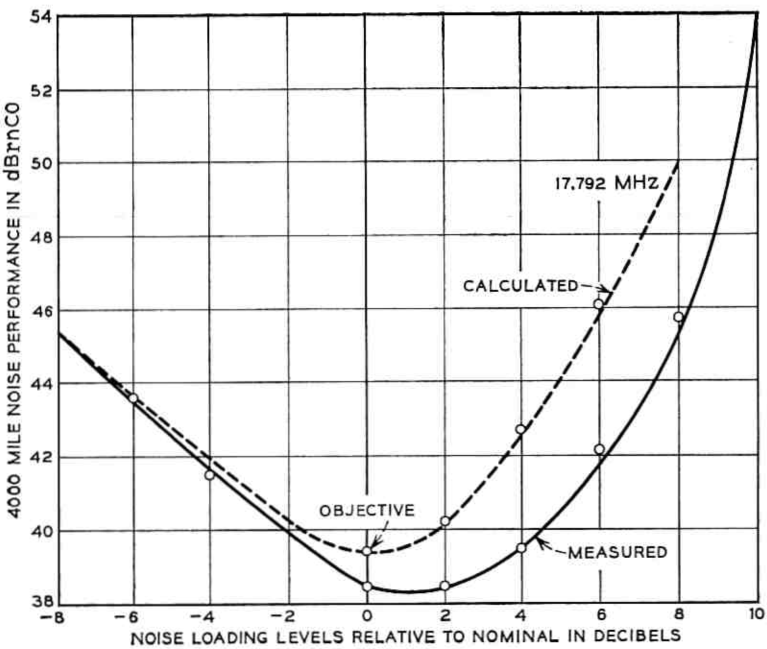

Fig. 7 — Noise performance extrapolated to 4,000 miles.

TABLE III —REPEATERED LINE LOAD CAPACITY

Required busy hour, fully misaligned rms load capacity = 10 dBm

Overlord criteria	Modulation	Differential gain	Impulse noise
Objective	$\Delta M_3 = 0.5$ dB	0.003 dB per repeater	1 count per minute 59 dBrnC0
Measured load capacity limit (dBm)	13	11	10
Load capacity margin (dB)	3	1	0

like impulses, sometimes referred to as "popping," are most severe when main station sections are operated at dc line voltages that approach the maximum of 1,800 volts. Under this operating condition, impulse noise performance is marginal when compared with the appropriate objective. However, only a relatively few such main station sections, approaching the maximum permissible spacing of 150 miles, would be encountered in a 4,000-mile system. Thus, a statistical advantage resulting from shorter average main station spacings and lower average operating line voltages, insures significant impulse noise margin in long haul circuits. Since the impulse noise magnitude in any given channel in the L-4 spectrum is sensitive to the line frequency of that channel, frogging rules which require transposition of mastergroups at maximum 800-mile spacings will help impulse noise margins by averaging the effects of impulse noise as a channel is frogged at specified intervals along the route.

6.3 *Repeater Load Capacity*

The line repeater load-carrying capacity has been evaluated with respect to three criteria: modulation, differential gain, and impulse noise. Measurements based on both the Ohio field trial and the first L-4 commercial installation indicate that the line repeaters provide sufficient peak power capacity to meet the requirements imposed by each of the three criteria with an overload margin which varies from 0 to 3 dB. Impulse noise performance appears to be most limiting, with essentially 0 dB of load carrying margin for a fully loaded, maximally misaligned 4,000-mile system. The load capacity margins observed for all three basic criteria are summarized in Table III.

6.4 *Crosstalk*

Repeatered line far end crosstalk shows a margin of approximately 3 dB versus the appropriate objective. This result reflects a substantial improvement over the results of the Ohio field trial, as a consequence of basic repeater design improvements instituted on the basis of field trial tests. Near end crosstalk performance is also satisfactory.

6.5 *Transient Response*

Transient response measured on 48 tandem regulating repeaters is well damped, with no observable tendency to spurious response or oscillation. Gain enhancement, with the pilot tone-modulated over a range of 0.01 to 1000 Hz, is within expected limits.

6.6 *Misalignment and Equalization*

Equalization of main station sections to within ± 0.5 dB has been achieved, meeting the 150-mile objective. This result includes the effect of the band edge dynamic gain regulator previously discussed, which corrects for low-end misalignment variations caused principally by temperature sensitive elements in the line repeaters. Throughout the rest of the L-4 spectrum, there is adequate A and B equalizer range to correct for residual fixed misalignment from imperfect deviation equalization in the regulating repeaters.

VII. PHYSICAL DESIGN CONSIDERATIONS

This section discusses some of the broad considerations which went into the physical design of the L-4 system. Detailed descriptions of all the equipment is given in several companion papers.[4–7, 9, 10]

Full realization of the potential of the L-4 system depended upon the physical design of its many units. This was particularly true for the repeaters required to operate in a manhole environment not previously considered in the development of coaxial systems in the Bell System. Careful planning and design were also needed in the development of the connecting and terminating gear at underground, "hardened" main stations. Space in these buildings would always be at a premium; the heavy circuit concentrations of L-4 made efficient spatial design imperative.

7.1 *General Environment*

The environment new to the design of coaxial repeaters is their operation in manholes subject to flooding and possibly to the effects of nuclear attack. These two requirements had major impact on repeater design. Actually these problems had been faced before when the hardened, transcontinental L-3 system was planned. However, that was more than a decade after the system had been developed and placed in manufacture. Rather than redevelop the L-3 equipment to meet a new environment, underground buildings were designed to provide the needed environment for the repeaters. The buildings were made impervious to water, and shock mounts were provided for all equipment.

Manholes were chosen as the housings for all of the L-4 line repeater equipment because of the very substantial savings over the L-3 type of underground installation. Even a poured manhole is less expensive if complete waterproofing is not required. However, the newly developed prefabricated manhole that could be used at the smaller basic and regulating repeater sites had estimated costs far below any other construction method.

Two options were open in the design of repeaters to operate in such installations. Hermetically sealed repeaters could be developed. These always present manufacturing difficulties; even the best seals will at times be ineffective, in which case there usually is no warning of failure. The alternate method was chosen: mounting several unsealed units in gas-tight cases maintained at cable gas pressure. The gas pressure serves two purposes: it prevents moisture entrance through small defects which might develop in the seal, and it warns of major troubles through low pressure alarms.

Designing L-4 equipment to withstand the effects of nuclear attack posed many questions in the planning stages. Certain of the blast phenomena, high winds and thermal shock, would be of no consequence because of the system being underground. This, plus the concrete construction, would also nullify or greatly weaken some of the elements in nuclear fallout. Three areas of concern remained. Two could cause circuit outages or shorten component or device life: nuclear radiation and electromagnetic pulse. The other, ground shock, was a threat to the physical well-being of the equipment.

Since many new devices and components were to be used, their sensitivity to nuclear radiation could not be determined until long after design decisions had to be made. Because of this and the fact that the only real threat was via access or air openings, it was decided to rely on building design for protection. Consideration of the electromagnetic pulse problem led to the same conclusion. At the time of development planning, this phenomenon had not been completely characterized. Studies indicated that the usual precautions against lightning such as the lead cable sheath and shield wires over the cable, plus the metal apparatus cases, would suffice for the manholes. For the main stations, similar cable construction and overall building shields were considered adequate.

The approach to the third problem, ground shock, differed in the two environments, the manned main stations and the repeater manholes. For the manned main stations, consideration of all the types of equipment involved led to a generalized solution. Shock-mounted structures, where required by the over-pressures, would be used throughout and individual equipment would only need to meet nominal strength requirements.

Overall shock isolation did not seem economically feasible for the L-4 manholes. Therefore, the development of the manhole equipment was directed at making each repeater strong enough to withstand ground shocks other than those of direct hits.

Four other problems faced the physical designer. It was realized at an early development stage that the decision to power the line repeaters from points as far as 150 miles apart would result in severe insulation and safety requirements. To maintain the life integrity of the transistors, heat transfer mechanisms required the most careful study and implementation. A third and all-pervading consideration was reliability, not only of each device and component, but of every connection between them. Lastly, in every design decision, the operating frequencies had to be weighed; the upper limits were twice those of any previous carrier system.

7.2 *Hazardous Voltages*

Early in the preliminary development it was estimated that about one-half ampere of direct current at between 1,500 and 1,800 volts to ground would be needed to provide the power for the line repeaters. This combination posed two serious problems. Most important was a serious personnel hazard, especially in the line repeater installations. Here, under adverse conditions of crowded quarters and the possibility of standing water, even after pumping, each piece of equipment could be a lethal threat. Signs warning of hazardous voltages and care to be taken are helpful; but, in times of stress their message is often forgotten. It was therefore vital that in every step of physical design the possibility be absolutely minimized of anyone being able to contact a lethal voltage inadvertently. The high voltage was also a potential hazard to the equipment itself. It required most careful consideration of the circuit components and of the methods used to insulate the various parts of the equipment to withstand voltages much greater than the working values.

7.3 *Heat Considerations*

Heat generated in the amplifiers of a transmission system can have deleterious effects on the passive components. Most of all, the heat can severely shorten the life of the very devices which produce it. Every effort must be made to transfer the heat from the devices to the surrounding environment. For the electron tube L-3 system this meant devising means of reducing plate temperatures to tolerable levels. For the L-4 system it would mean extensive study and design innovations in order to achieve operating transistor junction temperatures that would permit the almost limitless life potential of these devices to be realized. Treatment of the heat problem would be different for main stations and for manhole repeater locations. In main stations, central office conditions apply. In manholes, however, radiation and convection effects are negligible. Solution of the problem would depend on the design of very efficient conductive paths from transistor junctions to the soil surrounding the manholes.

7.4 *Reliability*

Reliability for systems such as L-4 has particular significance. With repeaters spaced every two miles it has a vital relationship to reasonable maintenance costs. This will be even more important as our land systems of the future have even closer repeater spacings. In the planning stages of L-4 it did not appear that the ultimate reliabilities of submarine cable systems or satellites were justifiable. This was based on the practical maintainability of a land system when compared to an ocean or space system. In addition, the land system could be provided with switchable spare line sections economically to cover short-time outages. However, every effort would be made to provide the most reliable components and interconnections within economical design methods.

7.5 *Operating Frequency Considerations*

The last important physical design problem was that of the frequency range of L-4. Certain design considerations become obvious as the upper operating frequency is extended. Maintenance of coaxial integrity and more stringent shielding are immediately apparent. In addition the length of interconnections and the placement of units are vital. All of these factors and their relative importance had to be gauged in the physical design of the line repeaters and the main station equipment.

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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-

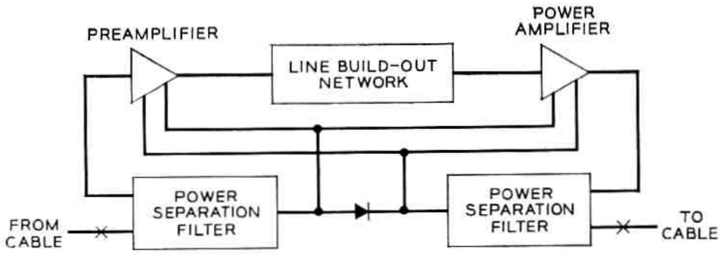

Fig. 1 — Simplified block diagram of basic repeater.

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-

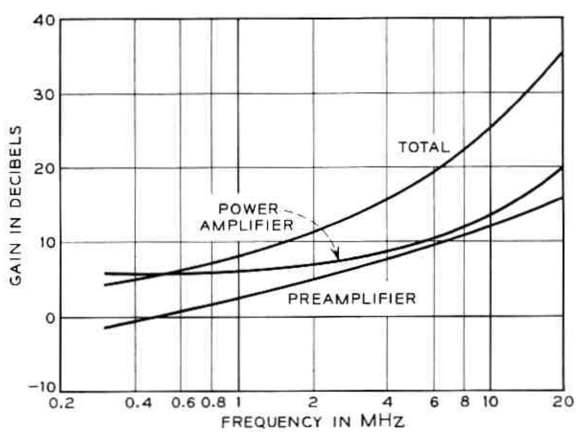

Fig. 2 — Contribution of preamplifier and power amplifier to total gain of basic repeater.

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.

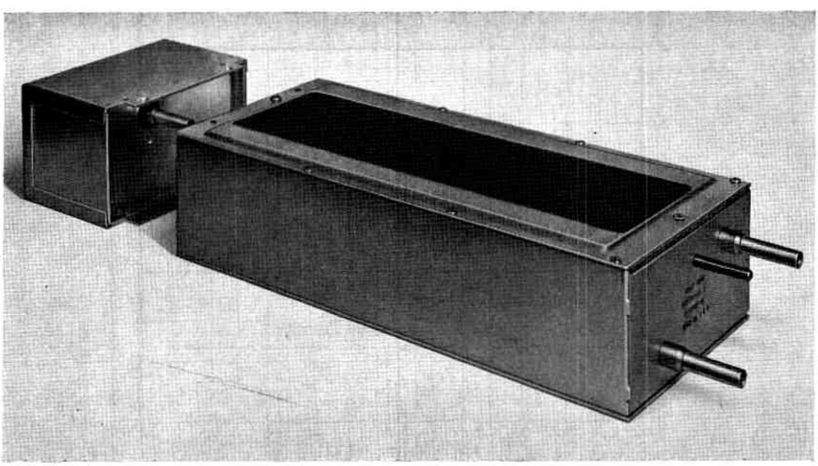

Fig. 3 — Basic repeater assembly with line build-out network (at left) detached.

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,

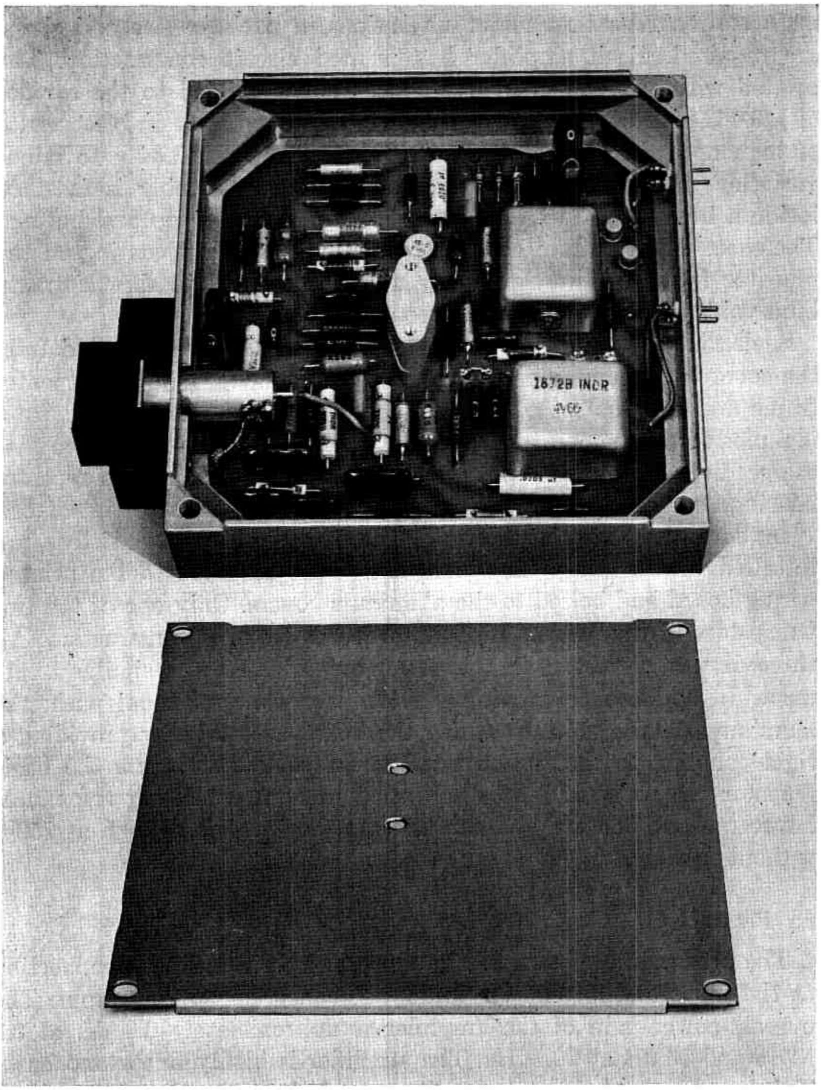

Fig. 4 — Top view of preamplifier with cover removed.

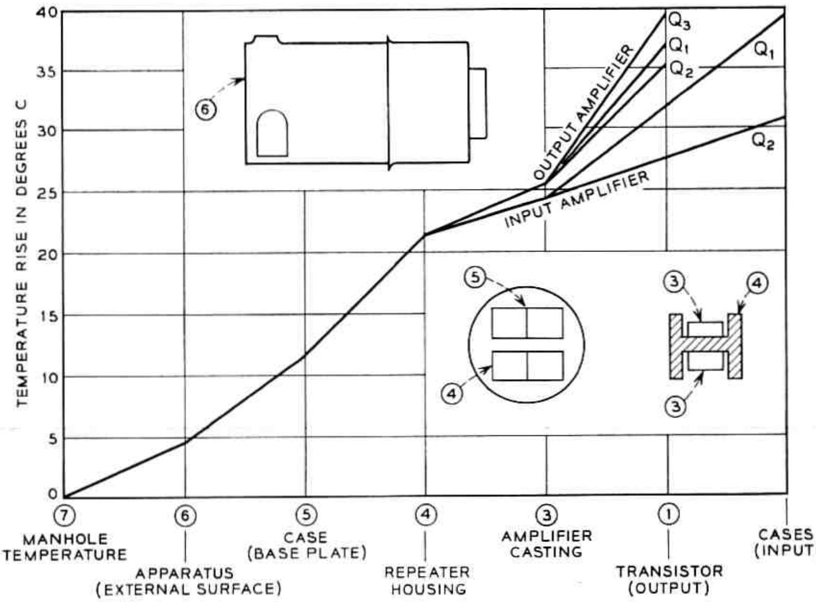

Fig. 5 — Temperature gradients for basic repeater installed in a manhole.

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,

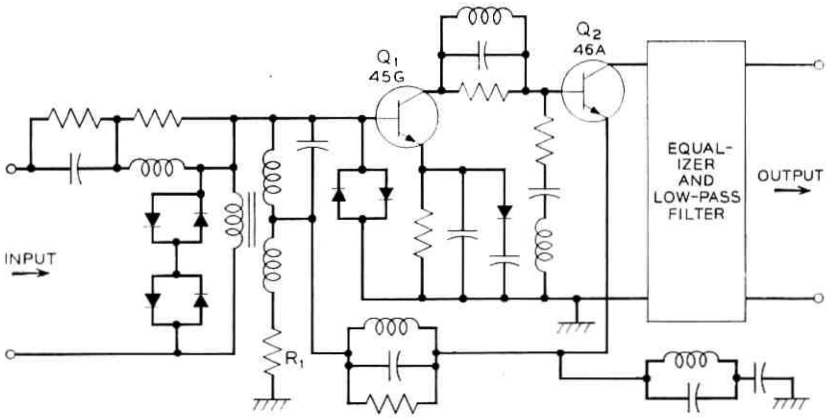

Fig. 6 — Schematic diagram of preamplifier (less bias circuitry).

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 $|\mu\beta|$.

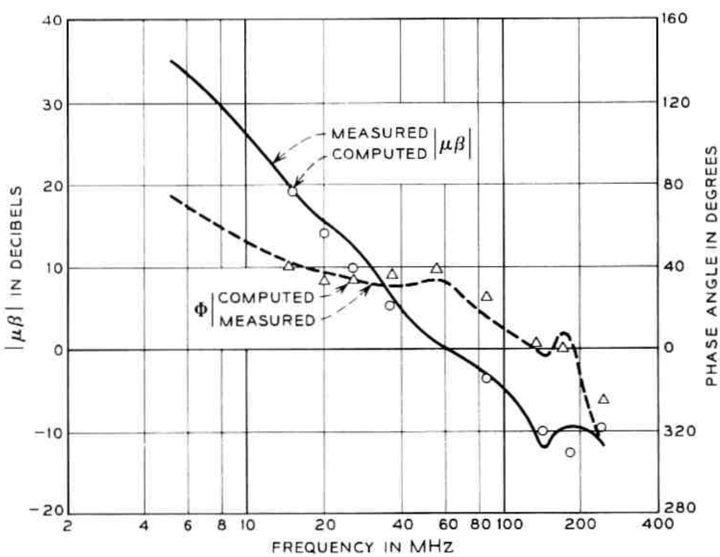

Fig. 7 — Typical measured and computed $\mu\beta$ or loop characteristic of preamplifier.

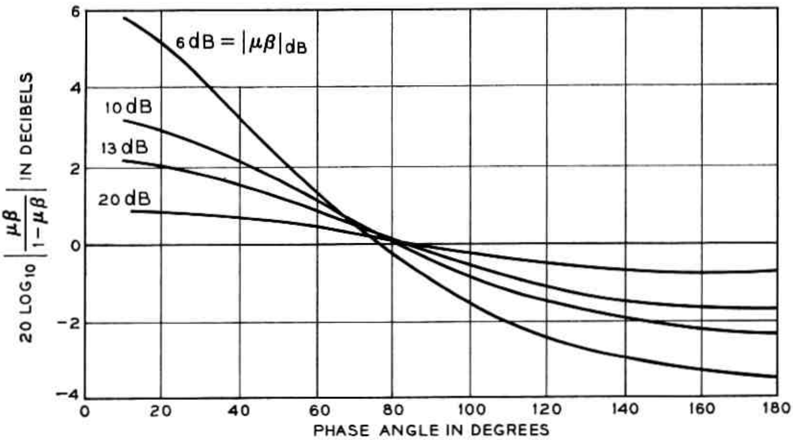

Fig. 8 — Calculated $\mu\beta$ effect vs $\mu\beta$ phase angle for various values of $|\mu\beta|$.

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,

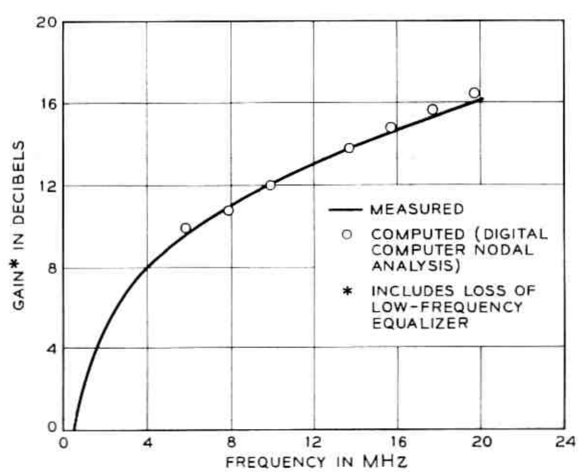

Fig. 9 — Measured and computed insertion gain of preamplifier.

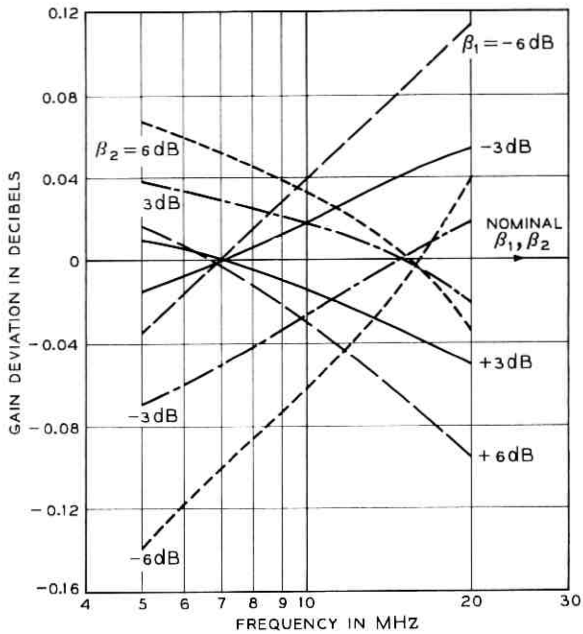

Fig. 10 — Calculated preamplifier gain deviations vs β of $Q1$ and $Q2$ transistors.

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

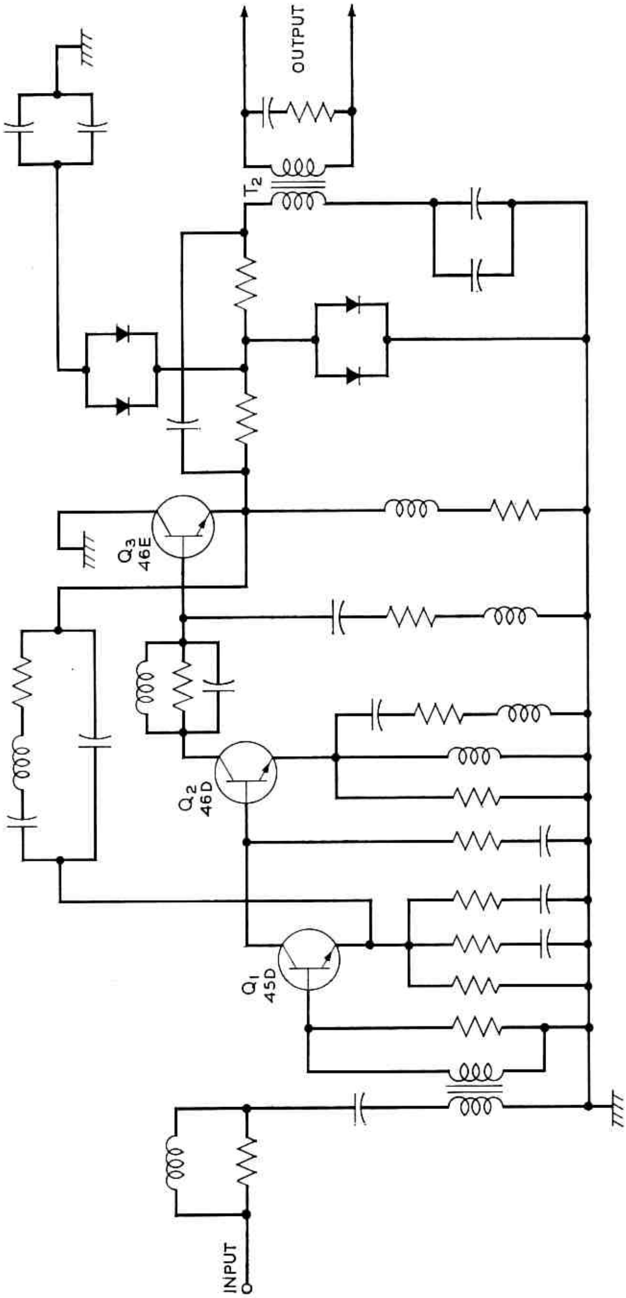

Fig. 11 — Schematic diagram of power amplifier (less bias circuitry).

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,

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

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

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

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

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.

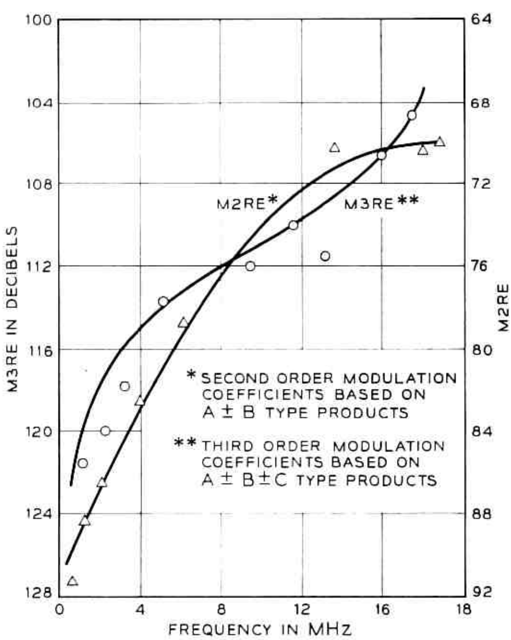

Fig. 12 — Second and third order modulation coefficients for power amplifier.

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

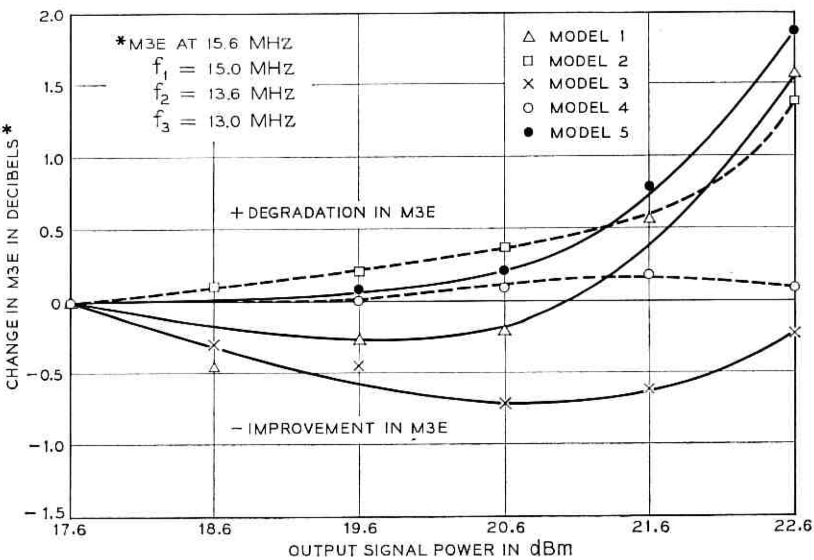

Fig. 13 — Change in third order modulation coefficient vs output signal power for power amplifier.

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;

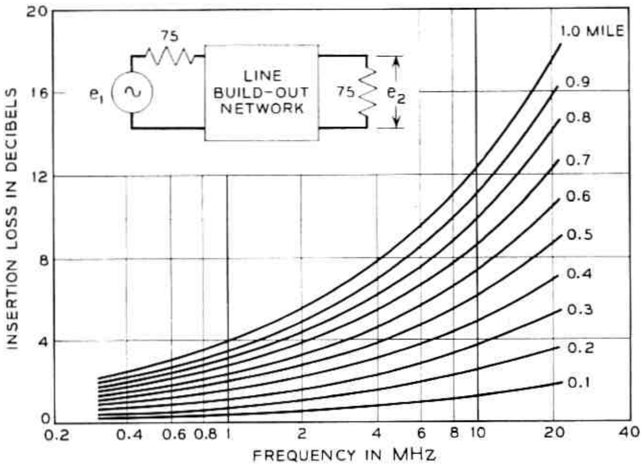

Fig. 14 — Insertion loss characteristics of line build-out networks (0.1 to 1.0 mile).

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.

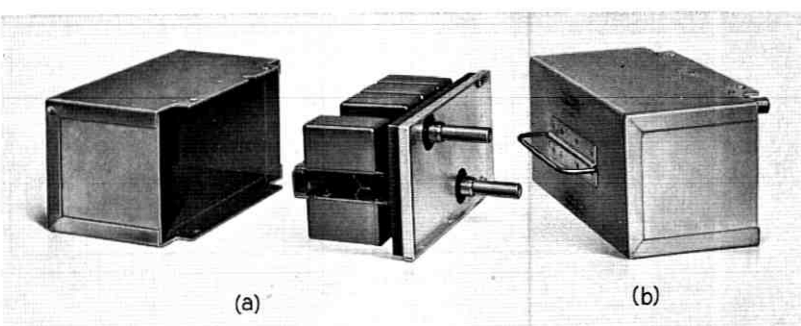

Fig. 15 — Mechanical construction of line build-out network: (a) network with cover removed; (b) completely assembled network.

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

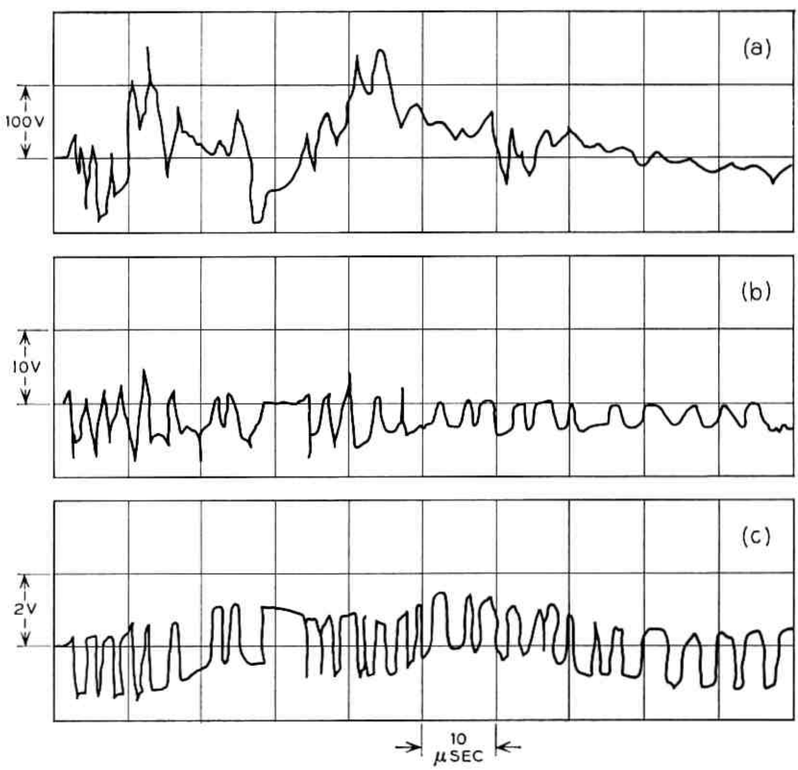

Fig. 16 — Transient voltages across (a) basic repeater input caused by a momentary short circuit applied to the high voltage line, (b) preamplifier primary protection diodes, (c) preamplifier secondary protection diodes.

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 ±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.

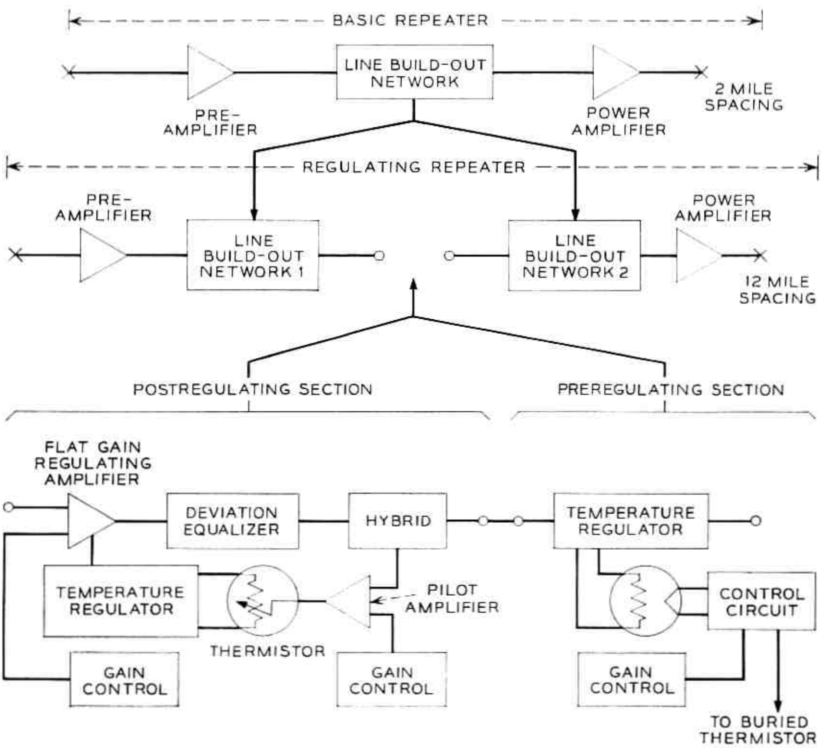

Fig. 17 — Simplified block diagram of regulating repeater showing build up from basic repeater.

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 $\pm 20°$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

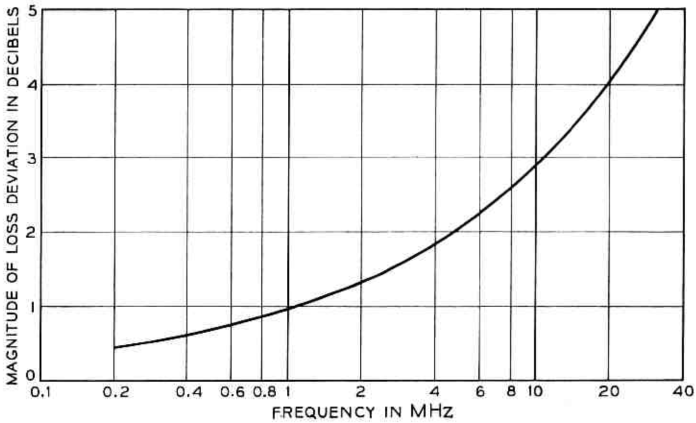

Fig. 18 — Magnitude of loss deviation resulting from an 18°F temperature change from nominal for 12 miles of coax 20.

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

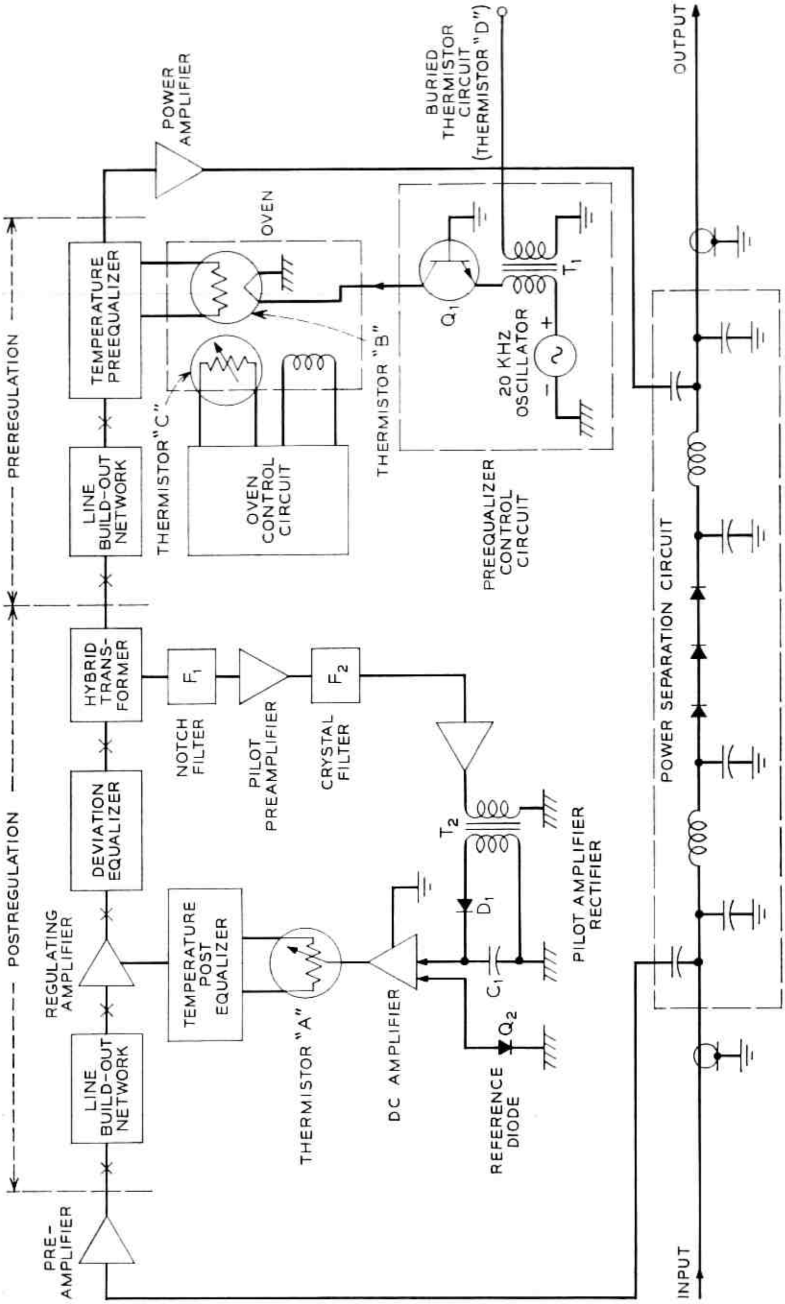

Fig. 19 — Complete block diagram of regulating repeater.

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

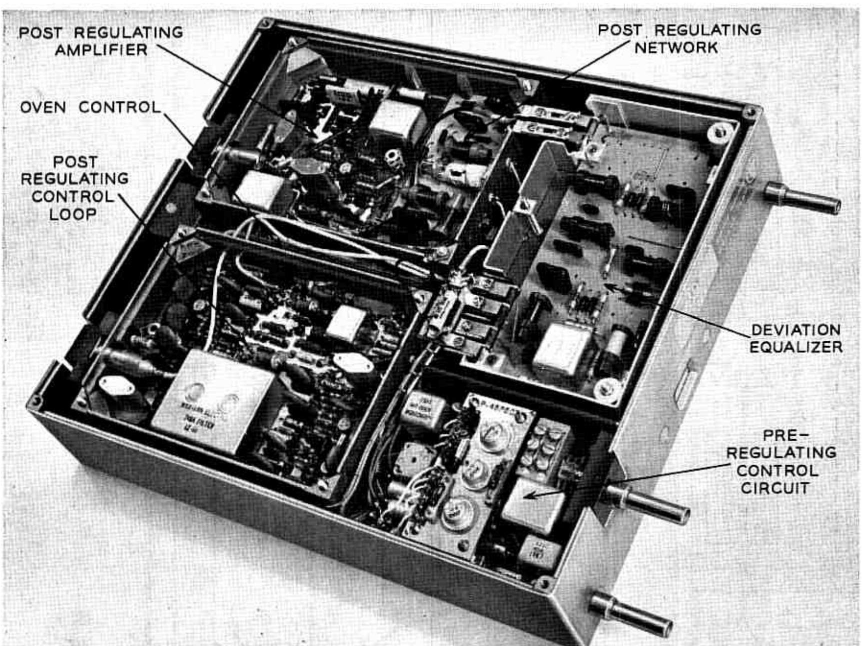

Fig. 20 — Top view of regulating repeater with cover removed.

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.

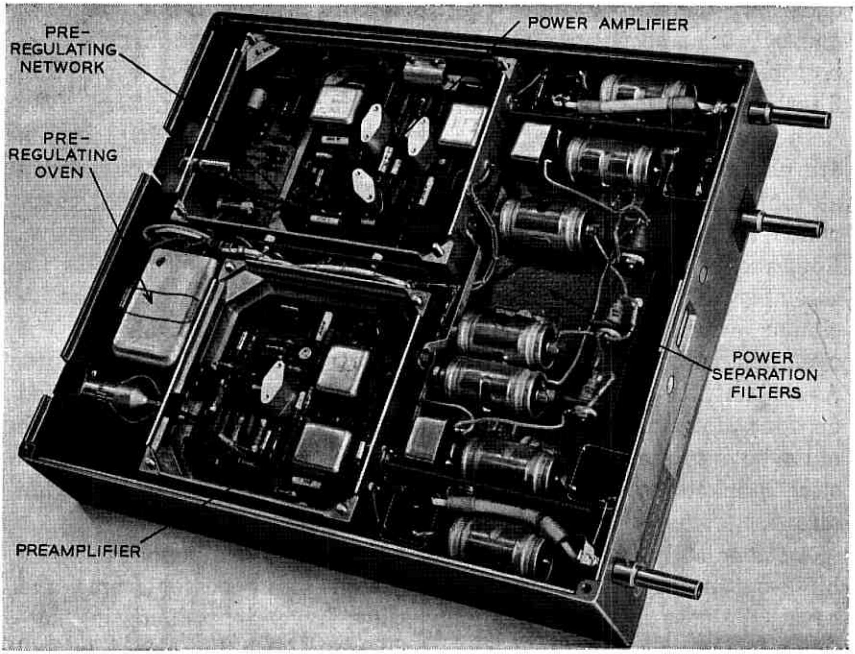

Fig. 21 — Bottom view of regulating repeater with cover removed.

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.

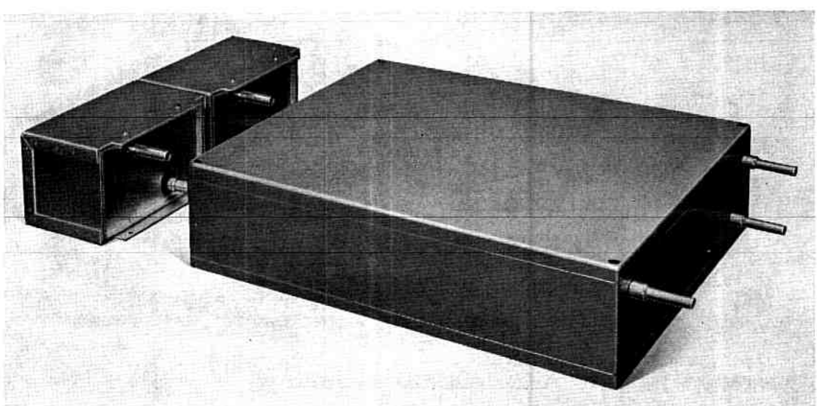

Fig. 22 — Regulating repeater assembly with two line build-out networks detached.

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

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

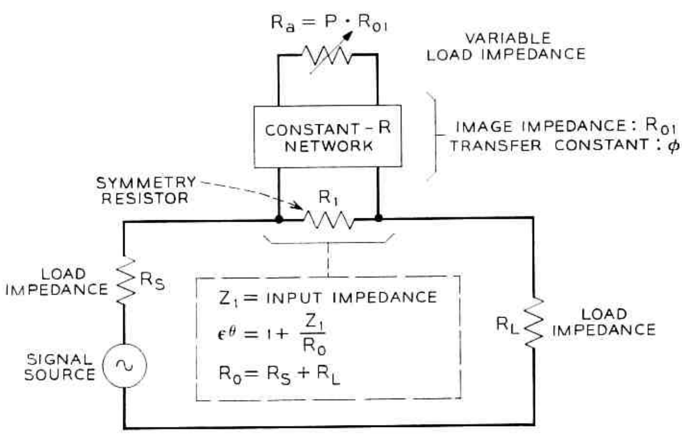

Fig. 23 — Series-type Bode regulator network.

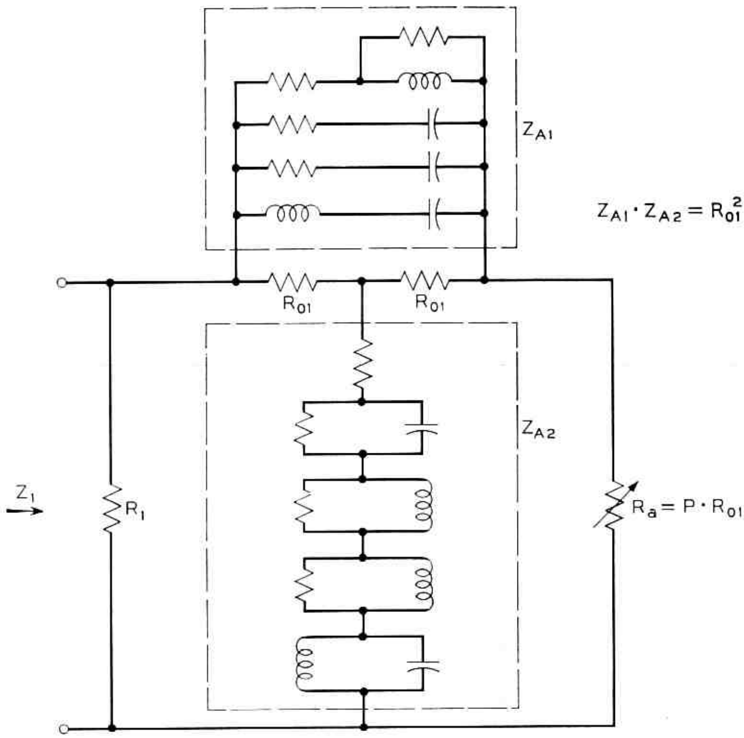

Fig. 24 — Preregulator network configuration.

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:

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

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

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.)

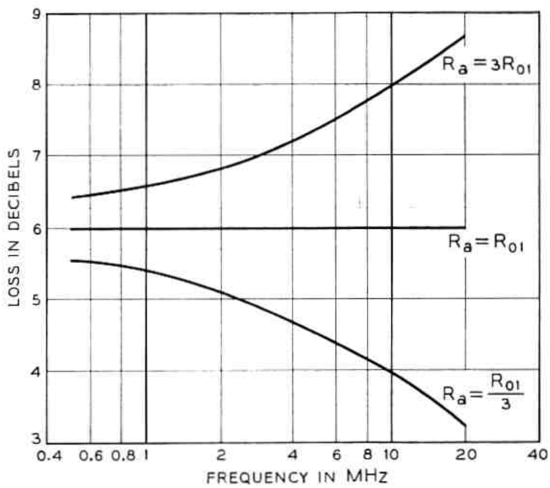

Fig. 25 — Insertion loss characteristic of preregulator network.

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

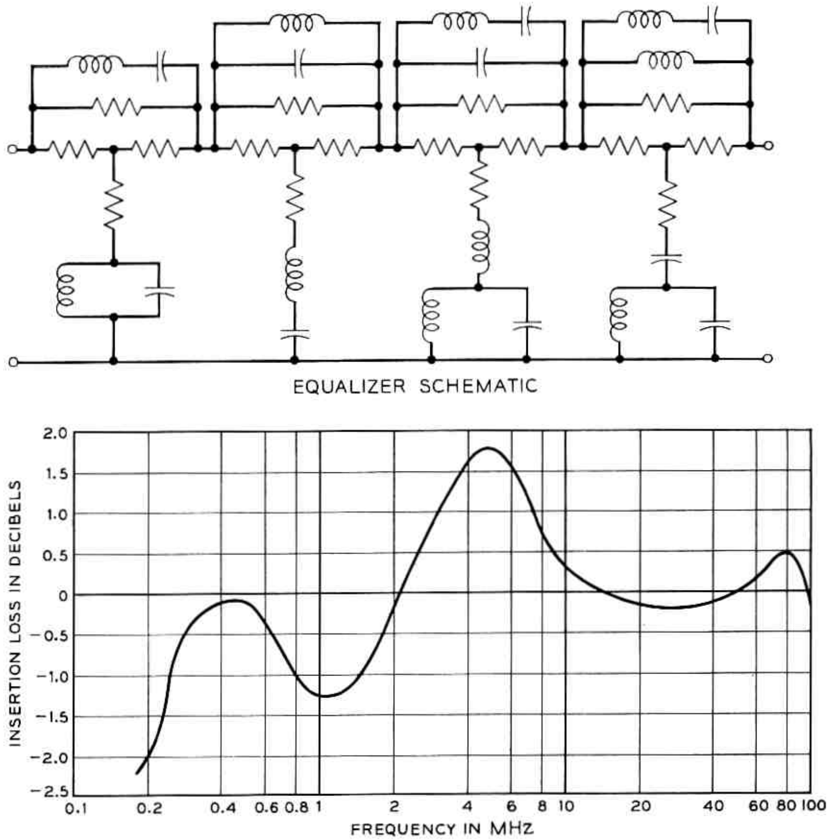

Fig. 26 — Deviation equalizer schematic diagram and requirements for a five-repeater regulating section.

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.

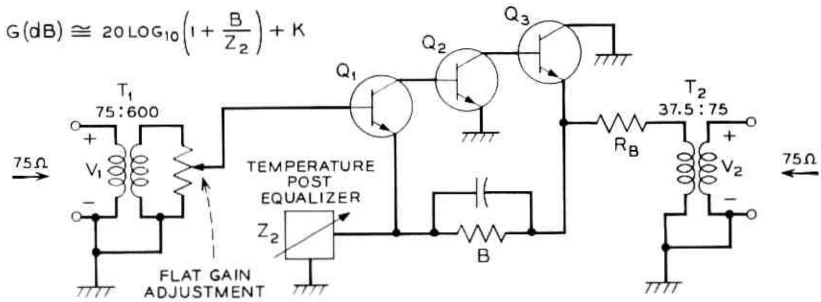

Fig. 27 — Simplified schematic diagram of regulating amplifier.

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

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

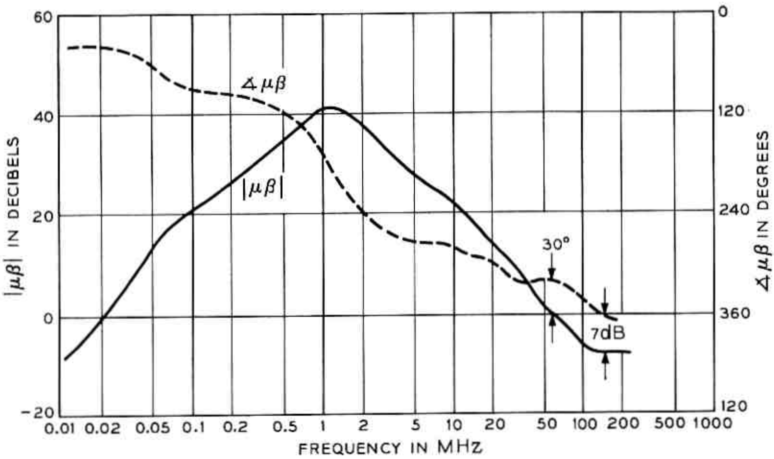

Fig. 28 — Typical $\mu\beta$, or loop characteristic, of regulating amplifier.

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

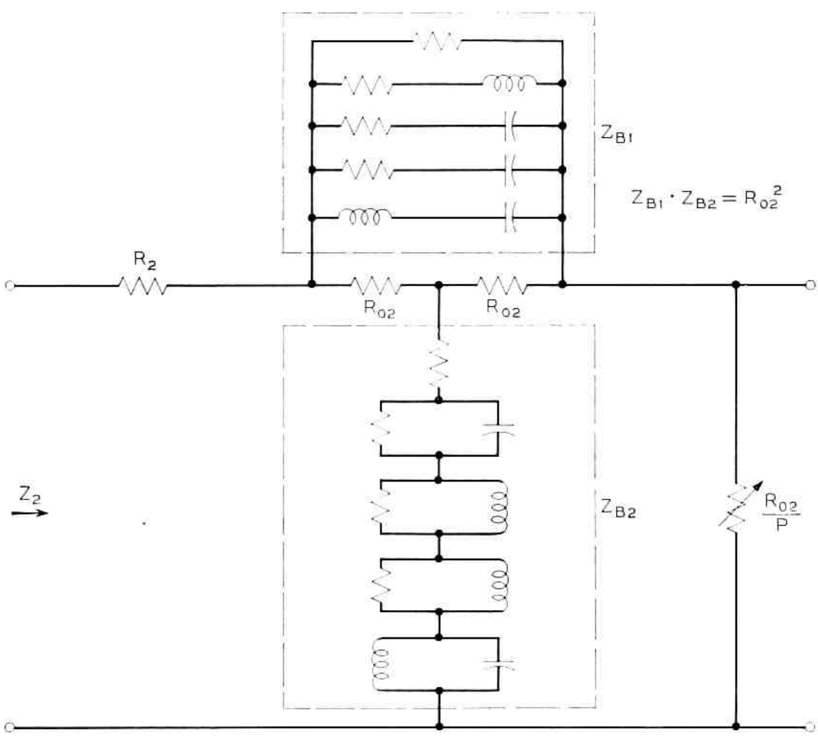

Fig. 29 — Schematic diagram of postregulator network.

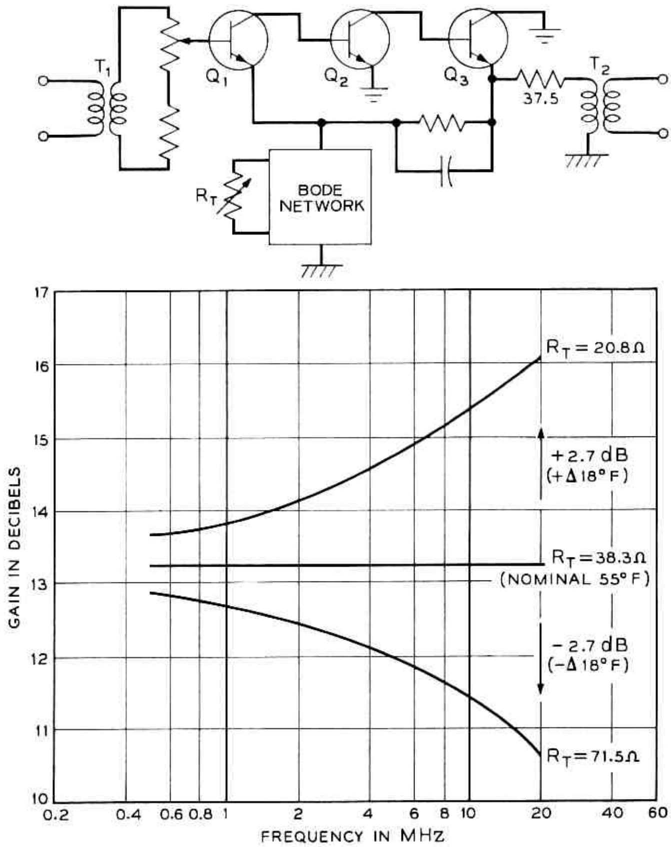

Fig. 30 — Performance characteristics of postregulator for a typical cable temperature swing.

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

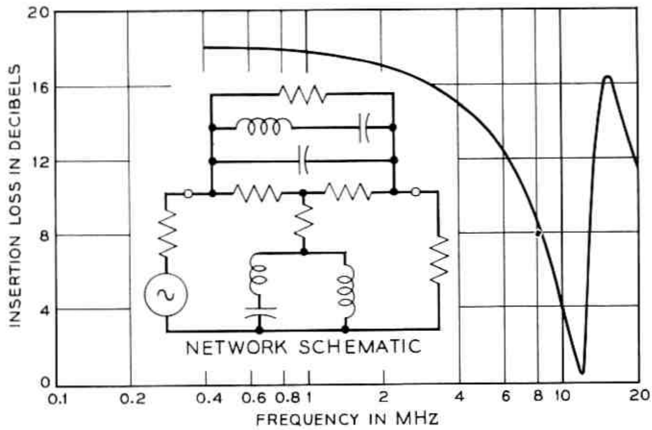

Fig. 31 — Insertion loss characteristic and schematic diagram of 11.648 MHz notch filter.

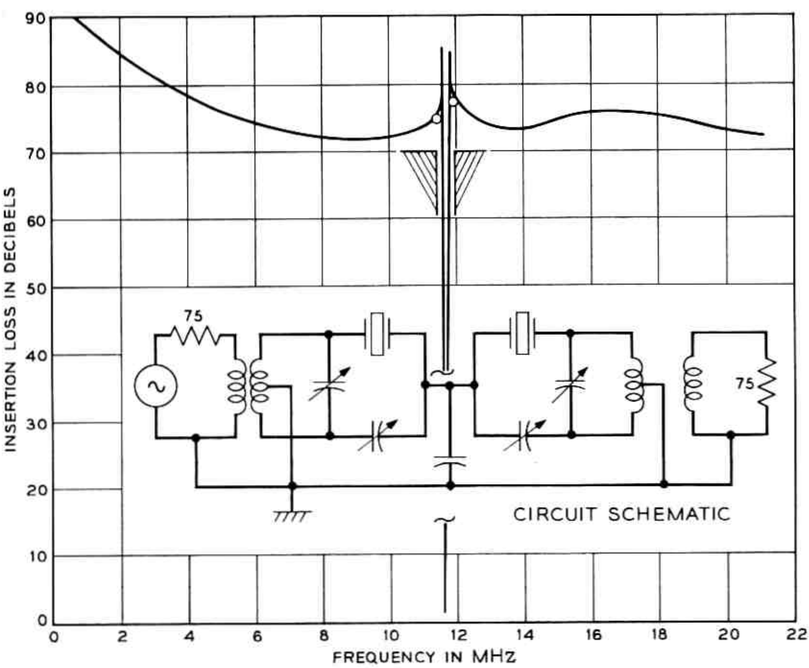

Fig. 32 — Insertion loss characteristic and schematic diagram of 11.648 MHz crystal pilot pick-off filter.

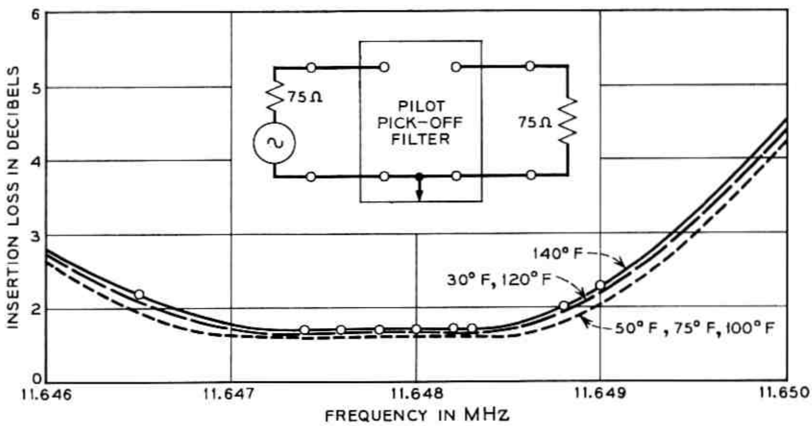

Fig. 33 — Inband temperature performance of pilot pick-off filter.

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.

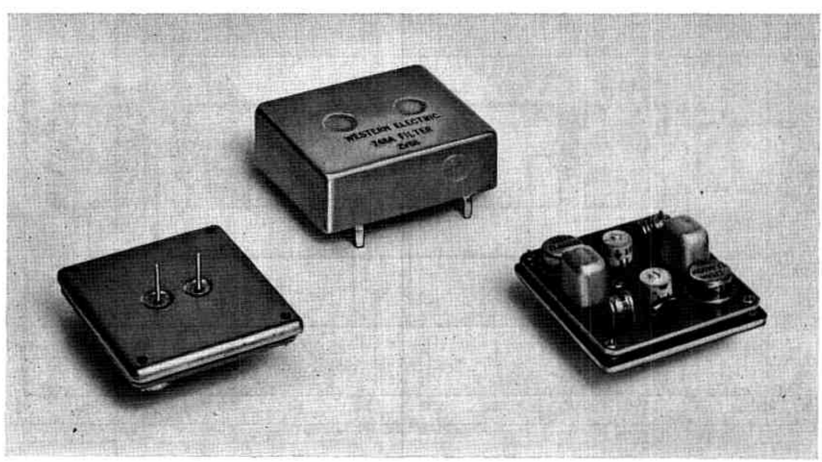

Fig. 34 — Mechanical construction of pilot pick-off filter.

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

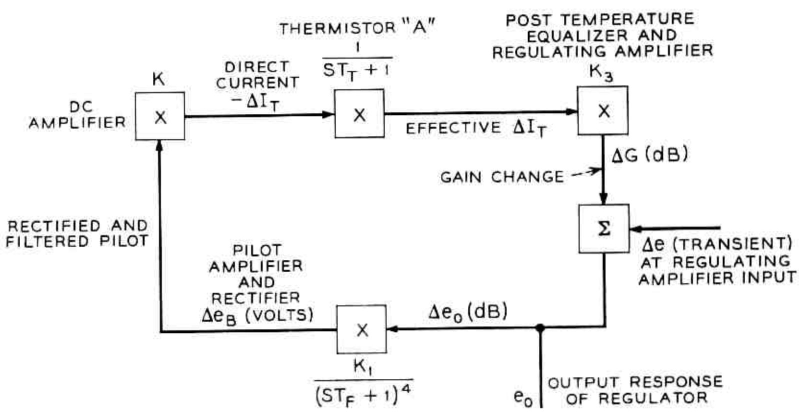

Fig. 35 — Mathematical model of the pilot-controlled postregulator.

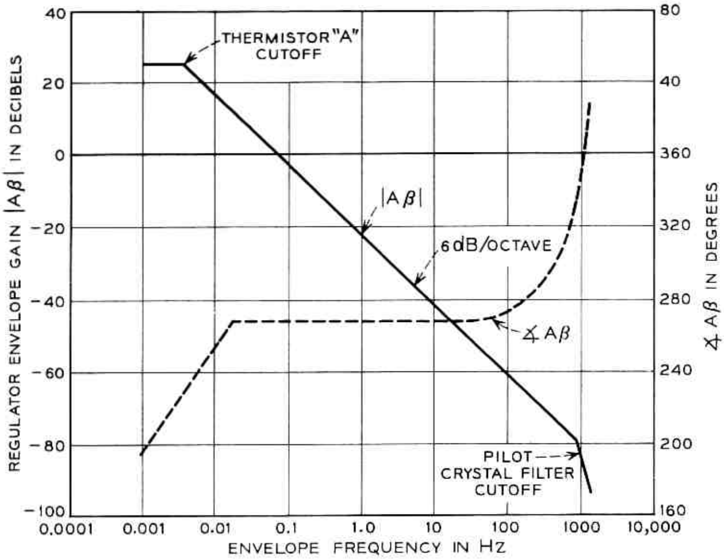

Fig. 36 — Measured postregulator envelope feedback characteristic.

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

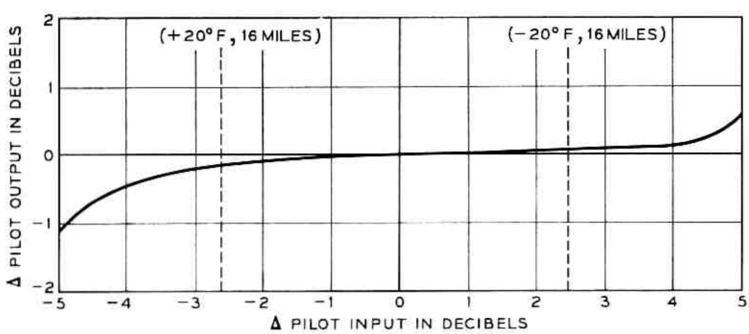

Fig. 37 — Regulating performance of pilot-controlled postregulator.

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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Equalizing and Main Station Repeaters

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The equalizing and main station repeaters, together with the L-4 control center and command looping circuits, provide for the equalization and remote fault location in the L-4 system. The use of pre- and postequalization procedures requires virtually identical circuits at the sending and receiving main station repeaters, while abbreviated versions of the same equipment are placed in up to two line equalizing repeaters per main section. In most cases these equalizing stations occur at approximately every twenty-fifth repeater in a completed system.

The fault-locating circuits are distributed over the entire system and the control circuits required for their activation are located at the equalizing points.

I. INTRODUCTION

The equalizing repeater is the most complex of the family of L-4 line repeaters; at least one is required in any line section longer than about 50 miles. A maximum of two equalizing repeaters can be placed between main station repeaters, which have a maximum separation of about 150 miles. The equalizing repeater contains all of the circuits found in a regulating repeater and thus provides, in part, the same features: the automatic gain correction required to track the cable loss variations with temperature and the fixed equalization required to compensate for the systematic or average component of the line repeater design error.

The main additional function of the equalizing repeater is the adjustable loss required to compensate for the random component of design error, for the effect of temperature on the many line repeaters, and for the effects of aging. This is accomplished in the equalizing repeater by the A equalizer which includes six independently adjustable Bode-type equalizer networks whose several adjustable loss characteristics affect different parts of the L-4 transmission band. The loss settings of each equalizer network are established remotely from the L-4 control center (located at a nearby main station) and are maintained thereafter by the solid-state memory circuits which are a part of the repeater.[1] Figure 1 is a block diagram of the repeater.

The equalizing portion of the main station repeaters includes an A equalizer and a B equalizer. The two are similar but the B equalizer includes ten different adjustable Bode-type equalizer networks which function in narrower frequency bands than the A equalizer networks. In that the A equalizer provides a relatively coarse correction of the system response, the B equalizer provides a correspondingly finer correction. As in the equalizing repeater, logic and memory circuitry are required to permit the remote control of the equalizer settings.

In the main station repeaters (Figs. 2 and 3), the functions normally found in a regulating repeater are found in part, in the line transmitting repeater at the transmitting main station and in part in the line receiving repeater at the receiving main station. In conjunction with the pre-emphasis networks of the terminal equipment, this arrangement provides the desired level shaping to the signal as it is applied to the coaxial line. The transmitting portion includes the cable temperature preregulation while the receiving portion includes the cable temperature postregulation.[2]

The receiving main station repeater also includes a band-edge regulator which inserts pilot-actuated gain correction affecting the region below 3 MHz. The shape of this automatically applied gain correction is like that of the lowest frequency A equalizer shape and is intended to provide dynamic equalization of that part of the band during the interval between regulator equalizer adjustments.

The L-4 fault-locating system consists of monitoring oscillators located in each line repeater apparatus case; remotely controlled power supplies, switching, and logic circuits located at each equalizing station; and the control center, located at main stations. Each equalizing and main station repeater includes a dc-dc converter to to provide the power necessary to energize a series-connected group of the monitoring oscillators; a switching circuit to select which of the two groups of oscillators adjacent to the equalizing point shall be energized; and logic circuits to decode the commands transmitted from the control center. The fault-locating system as a whole is discussed in detail in Ref. 1.

To prevent fault-locating operations in a given control section (which may consist of up to two main sections) from interfering with like operations in other sections, the receiving main station re-

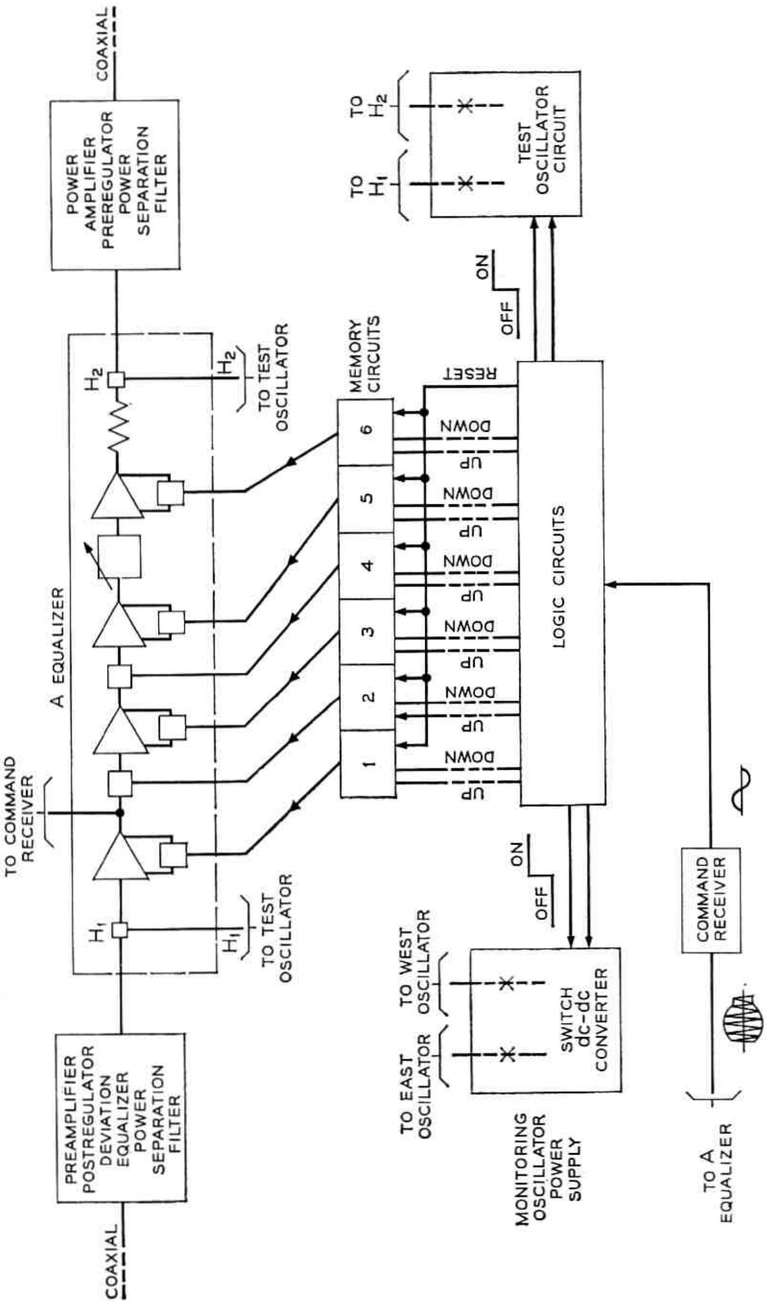

Fig. 1 — Functional block diagram of the equalizing repeater.

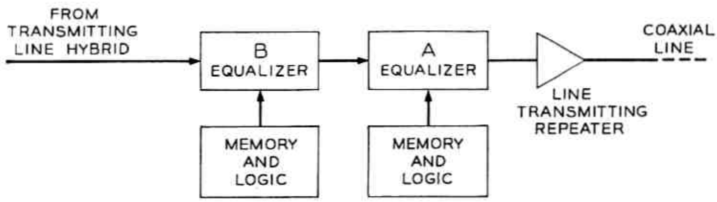

Fig. 2 — Block schematic diagram of the transmitting main station repeater.

peater provides for blocking the monitoring oscillator signals in the monitoring tone blocking circuit. In cases of two section control, this function is not supplied at the intermediate or power-feed station.

II. GENERAL FACTORS IN THE DESIGN OF THE L-4 EQUALIZING AND MAIN STATION REPEATERS

The equalization of the L-4 system involves three distinguishing characteristics:

(*i*) Bump shapes are used throughout.

(*ii*) Equalizer adjustments are made while the circuits are in service.

(*iii*) Line response is pre- and postequalized.

Considering the last of these features first, the use of partial pre-equalization and partial postequalization will, for a given response deviation, lessen the magnitude of the signal level deviation from nominal. If the nominal levels have been selected so as to minimize the loaded noise performance of the system, then any departure from these levels can only increase total system noise. Consequently, pre-

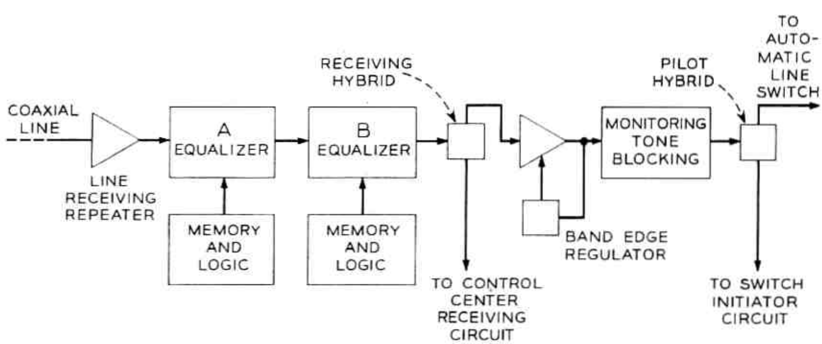

Fig. 3 — Block schematic diagram of the receiving main station repeater.

equalization can result in improved noise performance of the system. In systems limited by the load carrying capability of the repeaters, this feature can be used either to maximize the spacing of adjustable equalizers, to reduce the load requirements on the repeaters, or both.

In a system like L-4 where the noise performance is limited by modulation distortion and where the load capacity of the repeaters is not excessive, advantages in both areas are realized as a result of pre- and postequalization.

Given a section of modulation-limited system for which the accumulated misalignment is M_s (dB), and for which the nominal signal levels have been optimized with respect to total noise, the noise incurred in that section is increased by an amount determined by the method of equalization and by the relative importance of second- and third-order modulation distortion. If the total noise of the section is determined primarily by thermal and second order noise, the nominal system levels are selected so that the contribution from each of these sources is the same. Minimizing the noise in the misaligned condition requires that this relationship be maintained or, in other words, that both the thermal noise and second order noise incur equal penalties. This result is achieved if the signal levels are adjusted at the input to the section by $(M_s/2)$ dB, that is, if pre- and postequalization is applied in equal parts.

In general, the over-all penalty incurred in a section in which the noise is determined primarily by thermal noise and either voltage-adding or power-adding third order modulation noise will not be minimized by pre- or postequalization in equal parts. It can be shown, however (for moderate misalignment—up to about ± 5 dB), that pre- and postequalizing in equal parts incurs penalties which are trivially different from optimum.[4] Thus for sections in which the misalignment is in this range, the noise performance may be reoptimized by equal parts of pre- and postequalization whether the performance is limited by second order modulation distortion or by third order distortion.

This is the principle applied in the equalization of the L-4 response which, except in some cases for the lowest frequency supergroups, is usually misaligned between equalizing stations by considerably less than ± 5 dB. In fact, it will be later noted that the maximum adjustable range of the equalizer shapes is on the order of ± 4 dB and in most cases provides about a 2:1 ratio margin against anticipated needs.

The provision of accurate pre-equalization of relatively complex deviations requires a reverse channel or equivalent for the feedback of information on which the pre-equalizer settings may be based.

Fully adaptive equalization would require either the full-time surveillance of the channel to be equalized and full-time use of the reverse channel or a computer-like control system which would sample the channel at programmed intervals, and over the reverse channel would cause corresponding corrections in the equalizer settings.

In the latter case some kind of memory must be used to maintain the equalizer settings between samples and corrections. The L-4 equalization scheme corresponds in principle to the case in which the computer-like control system would sample the channel at programmed intervals, but in the present form cannot be programmed to sample automatically the coaxial line and cause the necessary corrections. The reverse channel in this case is the command band of a coaxial line transmitting in the opposite direction from the line being equalized.[1, 3] This single reverse channel is shared by all of the lines (up to 10 in the L-4 system) transmitting in the direction being equalized, as shown schematically in Fig. 4. The sampling and decision mechanism is found at the L-4 control center. Solid-state memory circuits maintain the equalizer settings between adjustments.

If the equalization procedures are to be implemented while the circuits are loaded and in service, it follows that the process of measuring the line response to determine what corrections, if any, are needed must not adversely affect the circuits in use. This precludes,

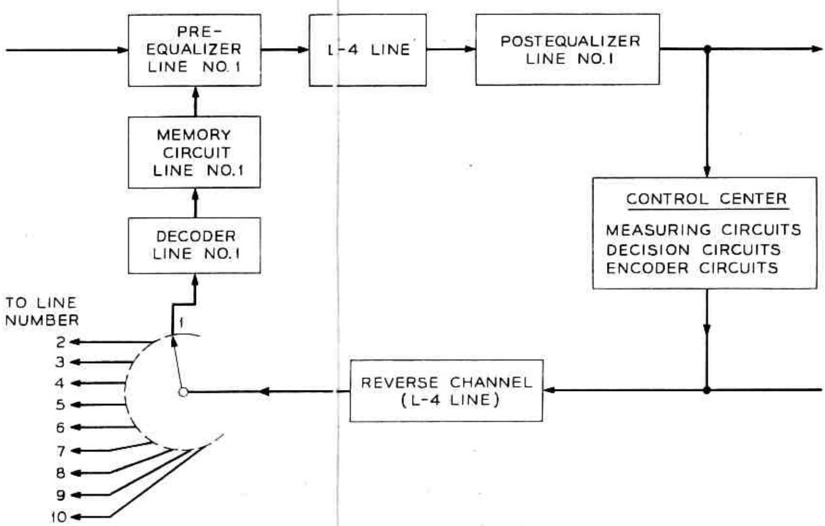

Fig. 4 — Functional schematic diagram of the L-4 equalizing system.

for example, using sweep techniques. However, the signals used to characterize system response must be separable from the message signals in the detection circuits. Consequently, the test signals associated with the adjustment of the A equalizers are located below mastergroup 1, between mastergroups and above mastergroup 6. As noted in Ref. 3, there is a 4 percent guard band between mastergroups. The test signals associated with the adjustment of the B equalizer are located above MG6 or within the mastergroups between the two submastergroups. Exceptions are the two lowest B test signals for which the test frequencies fall between adjacent supergroups of MG1. Guardbands between the submastergroups and supergroups are 56 and 8 kHz, respectively.

With this placement of the test signals across the transmission band, it is possible to realize practical filter designs which will in one case (control center detection process) separate the tones from the surrounding message power and in another case (at branching, dropping, and frogging points) block the test signals while passing the whole of the message spectrum.

The decision to use bump shape equalizers throughout the system is probably a bit more provocative than the other distinguishing features of the plan so far discussed. Obviously any of several approaches could provide the desired ultimate system response if the scheme were to use sufficiently complicated networks and adjustable shapes—and enough of them. The L-3 system, for example, uses 6 dynamic equalizers (each pilot-actuated and cause-associated), 15 cosine equalizers (including a flat term), and 5 bump equalizers (primarily to eliminate any unwanted effects of the cosine equalizer at the pilot frequencies of the dynamic equalizers). Bump shapes can be achieved by relatively simple Bode equalizer sections and offer attractive advantages over cosines with respect to realization and ease of adjustment. These features become particularly important in view of the desire to use both preequalization and in-service adjustment.

Preliminary calculations early in system development showed that a moderate number of bump shapes (between 10 and 20) would provide as good an equalized response as a corresponding number of cosine shapes. It further appeared that the bump shape approach would likely be superior in the final analysis when as many as possible of the bump shapes would be made cause-associated. Subsequent computer studies of the relative effectiveness of various combinations of adjustable shapes led to the selection of the family of 16 bumps shown in Fig. 5. Shapes 1–6 are located in the A equalizers while shapes 7–16 make up the B equalizer. At this time, however, only shape No. 1 may be characterized as cause-associated. As nearly as possible, this characteristic matches the loss variations with changing temperature of the many ferrite transformers located in the transmission path of the line repeaters.

III. EQUALIZING REPEATER

This section discusses the important factors involved in the design of the several distinct subassemblies which make up the equalizing repeater.

The regulating portion of an equalizing repeater is identical to the line regulating repeater except that access is provided to the output of the postregulating section and to the input of the preregulating section. This access is provided through high-voltage isolation transformers and permits the connection of the A equalizer between these points (Fig. 1).

3.1 A *Equalizer*

3.1.1 *General*

The A equalizer provides for the relatively coarse correction of the more complex deviations in system response which remain after the corrections applied by the regulating repeater. To the extent that the fixed deviation equalizers do not perfectly eliminate the systematic

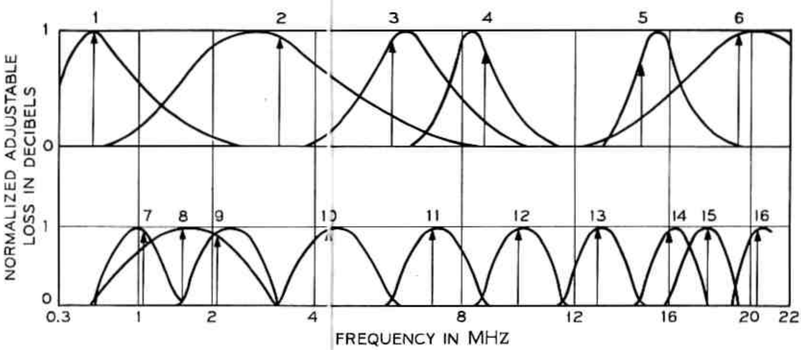

Fig. 5 — Normalized adjustable loss characteristics of the L-4 equalizer networks. The arrows denote the frequency of the test signals used as basis for adjustment.

repeater design error, the residual to be equalized will include a contribution from this source. To the extent that the latter becomes a significant part of the total residual, the resultant equalized response from main section to main section will tend to include like characteristics which will accumulate systematically along the system. It is therefore vital that the response to be compensated for by the equalizers include a minimum component of this sort.

The factors considered in Section II led to the specification of the equalizer shapes numbered 1–6 in Fig. 5 for the A equalizer. These are the relatively broad shapes required to make the necessary first-order correction which in turn permits the advantages with respect to signal-to-noise ratio and overload to be realized.

The main constituents of the A equalizer are shown in Fig. 6 and include four flat gain amplifiers, six adjustable Bode equalizer networks, and the elements and circuits required to permit remote control.

3.1.2 *Amplifier Design and Performance*

3.1.2.1 *General.* The equalizer amplifiers basically provide enough gain to compensate for the loss of the necessary adjustable equalizers. In addition, there is some loss associated with the circuits which permit both the coaxial line and the A equalizer to be interrogated as to response at the six A test frequencies. Finally, there is loss resulting from the trim equalizer network, which is required as a mop-up in order that the over-all A equalizer response in its reference state be as nearly as possible 0 dB throughout the L-4 band.

It is desirable, from considerations of noise and reliability, to minimize the number of amplifiers required to achieve this basic objective. This must also be balanced with the practicability of the resultant requirements on permissible noise, nonlinear distortion, and load capacity. A satisfactory compromise is the configuration of Fig. 6 wherein four of the adjustable Bode networks are incorporated in the feedback paths of four amplifiers, while the other two and the trim equalizer are connected between amplifiers.

Since the adjustable characteristics required of the A equalizer affect a fairly wide range of frequencies, it is particularly desirable that the amplifier configuration selected permit good isolation of the several equalizer networks to minimize interaction effects, and that this isolation be achieved with a minimum of loss. The configuration of Fig. 7 satisfies this requirement and is characterized by the following features:

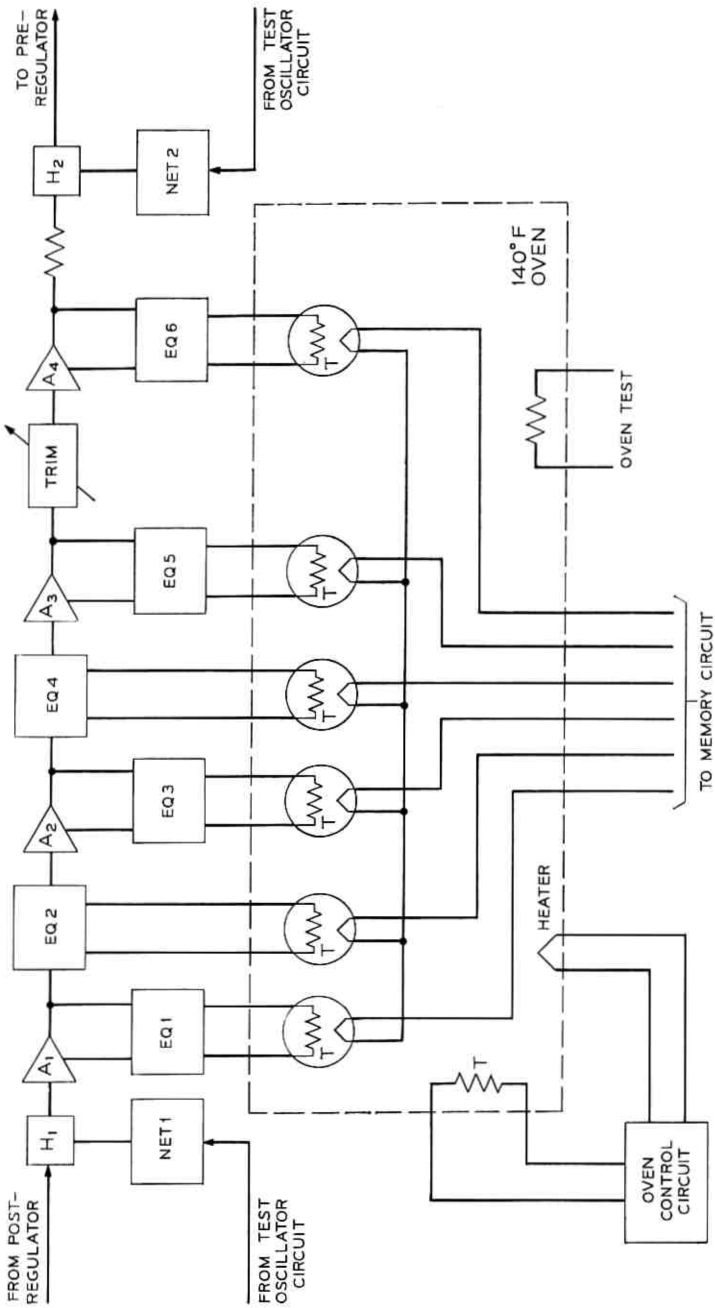

Fig. 6 — Block schematic diagram of the A equalizer. For clarity the two ovens and oven control circuits actually used are shown as one.

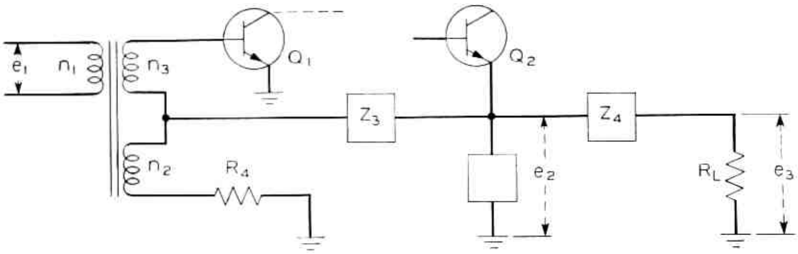

Fig. 7 — Simplified schematic diagram of the equalizer amplifier. Z_3 and Z_4 would normally be realized by the adjustable impedance of equalizer networks.

(*i*) The output impedance is approximately zero ohms and the voltage gain is essentially independent of load impedance.

(*ii*) The voltage gain is shown in the appendix to be

$$\frac{e_2}{e_1} \doteq \frac{n_3}{n_1}\left(1 + \frac{Z_3}{R_0}\right)$$

where $R_0 = n_3/(n_2 + n_3)R_4$. For $n_3 = n_2 = n_1$,

$$\frac{e_2}{e_1} \doteq 1 + \frac{Z_3}{R_g},$$

where R_g is the nominal resistance terminating the primary winding and in this case is 75Ω.

The loss associated with an equalizer Z_4 inserted in series between the output transistor and the load resistor, R_L, as shown in Fig. 8, is

$$\frac{e_2}{e_3} = 1 + \frac{Z_4}{R_L} = 1 + \frac{Z_4}{R_g}, \quad \text{if} \quad R_L = R_g .$$

Where the range required of the adjustable networks is similar, it is possible to design a family of like networks which operate at identical impedances and which require the same range of variable control resistance. All of the adjustable shapes could thus be achieved with virtually the same network and thermistor design.

Where greater or lesser range is required, this can be readily achieved by the specification of a different input hybrid transformer, one which can continue to provide the necessary load impedance to any series-connected networks while adjusting the β-circuit impedance at which the feedback network operates.

In a similar fashion the flat gain of the configuration can be adjusted without affecting Bode network performance, if desired, by specifying appropriate turns ratios on the transformer.

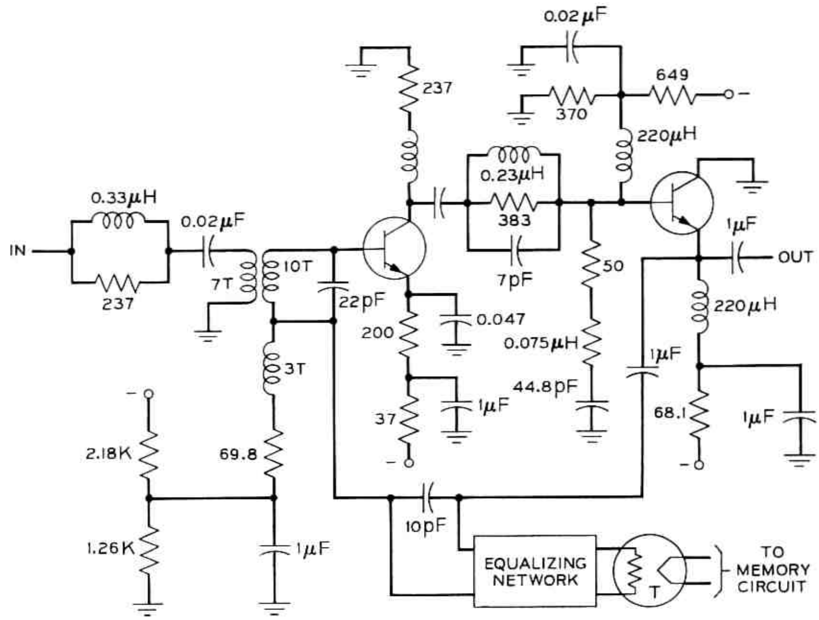

Fig. 8 — Schematic diagram of a typical equalizer amplifier. The circuit shown is that of A_4 in Fig. 6.

The configuration can be realized with a relatively simple two-stage design with common emitter input section and common collector output section, permitting excellent performance in respect to noise figure and linearity, respectively.

3.1.2.2 *Gain Considerations.* The overall voltage gain of the configuration of Fig. 7 is given by

$$\frac{e_3}{e_1} = \left(1 + \frac{Z_3}{R_g}\right)\left(1 + \frac{Z_4}{R_g}\right)^{-1},$$

where $R_0 = R_g$ and $n_3 = n_2 = n_1$; and $e_3/e_1 = 1$ if Z_3 and Z_4 assume equal values, which is very nearly the case for the Bode network designs used when each is in its reference state.

Consequently, a convenient building block of approximately zero dB gain (with the adjustable networks in the reference or "flat" state) can be readily achieved with this amplifier configuration by incorporating one equalizer network in the β-circuit, one equalizer network in series with the output transistor, and by specifying that $n_1 = n_2 =$

n_3. Under these conditions the equalizer networks will provide the same adjustable response capability whether connected in series with the output or connected in the β-circuit. As can be seen in Fig. 6 there are two such amplifier-network assemblies used in the A equalizer (A1–EQ2 and A2–EQ4).

The third amplifier of the A equalizer has an equalizer network only in the β-circuit and consequently provides a net voltage gain of

$$20 \log \left(1 + \frac{Z_3}{R_0}\right) \text{dB}.$$

As is described in subsequent sections, the value of Z_3 is determined, for operation with a particular R_0, according to the loss adjustment range required of the network and usually the wish to make that adjustment symmetric with respect to the reference or flat condition. In this case the value is approximately 93 ohms and, with the ability to vary the network control resistance between about 27 ohms and 665 ohms (using the indirectly heated thermistor), permits a loss variation at the network center frequency of about ± 4.0 to ± 4.5 dB. As a result, the gain of the amplifier A_3 of Fig. 6 is

$$20 \log \left(1 + \frac{93}{75}\right) = 7.0 \text{ dB},$$

and is approximately flat over the L-4 band.

The fourth amplifier in the string making up the A equalizer, A_4 of Fig. 6, is similar to A_3 in that the β-circuit alone includes an adjustable network. In this case it is the network having effect in the lowest regions of the L-4 message band and in the command and switching channels. There are two factors, however, which result in minor differences in its design. First, the adjustable range required of the low end equalizer network is greater than that required of any of the others—more nearly ± 6.0 dB than ± 4.0 dB. Second, some additional gain (beyond the 7 dB achieved from the circuit used in the other three amplifiers) is required to compensate for the total loss of the input and output connecting circuits and the trim equalizer. The flexibility of the selected configuration permits both of these goals to be achieved with a simple change of input transformer, all the while maintaining the standard over-all amplifier input impedance.

The same equalizer network design applies with an adjustment in the symmetry resistor to restore a symmetrical relationship between maximum and minimum loss and the reference loss. Selection of an input transformer for which $n_1 = 7$, $n_2 = 3$, and $n_3 = 10$, where

$$R_0 = \frac{n_3}{n_2 + n_3} R_4 = 49\Omega,$$

and

$$\frac{e_2}{e_1} = \frac{n_3}{n_1}\left(1 + \frac{Z_3}{R_0}\right) = 4.36,$$

provides a nominal gain of 12.8 dB with about ±6.0 dB adjustable range. The control resistance variation required by all of the networks is approximately 27 to 665 ohms, with reference or flat loss achieved when the network is terminated in 135 ohms. (A schematic for A_4 is shown in Fig. 8.)

3.1.2.3 *Noise Figure Considerations.* The noise figure of the amplifiers making up the A equalizer is not as significant to over-all system performance as that of some of the other line amplifiers, since A equalizer amplifiers are fewer and the transmission level within the A equalizer is not the lowest in the system. Nevertheless, it is necessary to minimize noise contributions from whatever source; a total A equalizer objective for noise figure less than 18 dB is satisfactory from the overall viewpoint. Fig. 9 shows the A equalizer constituents affecting noise figure and readily permits a satisfactory set of objectives for the individual amplifier noise figure to be developed. The symbol C in Fig. 9 is used to designate the negative of transmission level, or the transmission level expressed in dB below zero level.

The zero level noise contribution of the kth amplifier will be, in

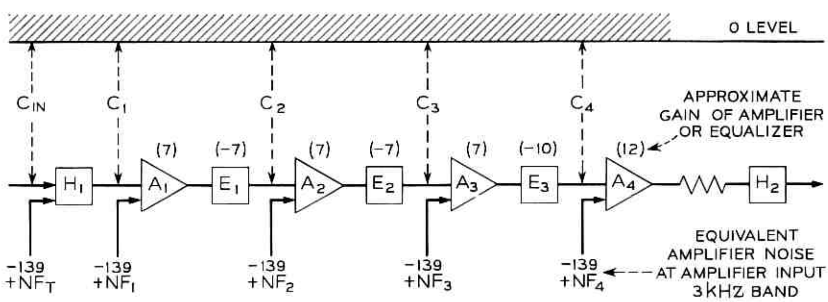

Fig. 9 — Block Diagram of the A equalizer for noise figure calculations. Indicated are the equivalent noise sources and the approximate amplifier and network gains.

a 3 kHz band:

$$P_k = -139 + NF_k + C_k \text{ dBm}.$$

The total noise contribution of the amplifiers making up the equalizer will be

$$P_{Nt} = \text{“}\sum_{k=1}^{4}\text{”} (-139 + NF_k + C_k) \text{ dBm} \tag{1}$$

where the “$\sum$” implies addition on a power basis.

The approximate nominal losses and gains are indicated on Fig. 9; it can be seen that, for these particular values,

$$C_1 \approx C_2 \approx C_3 \triangleq C.$$

If it is further assumed that $NF_1 = NF_2 = NF_3 = NF$ (which is not unreasonable, since the design of amplifiers 1 to 3 has been shown to be alike for other reasons) then (1) can be rewritten as:

$$\begin{aligned} P_{Nt} &= (-139 + NF + 10 \log 3 + C) \text{ “}+\text{” } (-139 + NF_4 + C_4) \\ &= -139 + [(NF + 10 \log 3 + C \text{ “}+\text{” } (NF_4 + C_4)]. \end{aligned} \tag{2}$$

It can be seen in Fig. 9 that C_4 is about 3 dB larger than C since the trimmer loss is about 3 dB larger than the loss of the series equalizer networks. If NF_4 can be realized at 3 dB lower than NF, then (2) can be simplified to

$$P_{Nt} = -139 + NF + 10 \log 4 + C \text{ dBm}. \tag{3}$$

In terms of the equivalent noise figure of the overall A equalizer,

$$P_{Nt} = -139 + NF_T + C_{IN}. \tag{4}$$

Equating P_{Nt} from (3) and (4) yields

$$NF_T = NF + 10 \log 4 + C - C_{IN}.$$

Since C–C_{IN} is just the loss of the input hybrid connecting circuit, L_H,

$$NF_T = NF + 10 \log 4 + L_H.$$

Thus

$$NF = NF_T - 10 \log 4 - L_H.$$

For $L_H = 1.5$ dB and letting $NF_T = 18$ dB,

$$NF = 18 - 6 - 1.5 = 10.5 \text{ dB}.$$

Recalling the assumption that the noise figure of the fourth amplifier would be 3 dB better than the others,

$$NF_4 = 10.5 - 3 = 7.5 \text{ dB}.$$

Consequently, satisfactory over-all noise figure performance may be achieved in the A equalizer if the configuration used in the first three slots realizes a noise figure of 10.5 dB or less and the configuration used in the fourth position, 7.5 dB or less. These are well within the capability of this configuration with the devices used.

The noise figure of the individual amplifier is determined primarily by the noise figure of the input transistor and the coupling loss of the input connection. Other considerations are the extent to which the input stage is mismatched with respect to noise figure and the extent of local feedback on the first stage.

With respect to local feedback, it can be shown that the effect of a resistor in the emitter of a common emitter transistor stage is entirely analogous to that of r_b' and that the effective noise figure of the stage, based on the equivalent circuit for noise shown in Fig. 10, becomes[5,6]

$$F \approx 1 + \frac{R_E}{R_g} + \frac{r_b'}{R_g} + \frac{r_e}{2R_g} + \frac{(1-\alpha_0)\left\{1 + \left[\frac{f}{(1-\alpha_0)^{\frac{1}{2}} f_\alpha}\right]^2 (R_g + R_E + r_b' + r_e)^2\right\}}{2\alpha_0 r_e R_g}. \tag{5}$$

(Apart from the loop gain considerations, an added factor favoring the common emitter input stage is the realization of the optimum

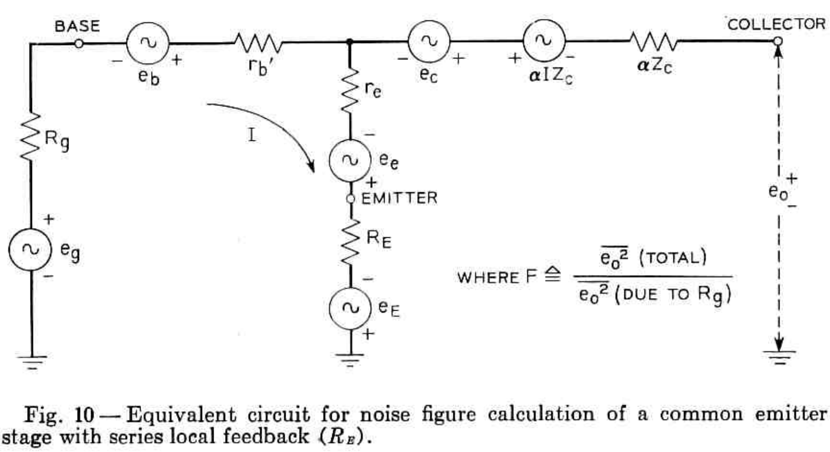

Fig. 10 — Equivalent circuit for noise figure calculation of a common emitter stage with series local feedback (R_E).

noise figure and maximum device gain at very nearly the same source impedance.)

Because of the lesser gain required of the amplifiers in the first three positions in the A equalizer, it is necessary and desirable in achieving a satisfactory loop gain response to apply series negative feedback to the input common emitter stage. From (5) it can be seen that this will degrade the effective noise figure of the stage to an extent which depends on the magnitude of R_E. Defining this degradation to be Δ_{RS} (in dB), the noise figure of the amplifier may be established by

$$NF = NF_{trans} + L_H + \Delta_{RS} , \qquad (6)$$

where L_H is the hybrid transformer coupling loss and NF_{trans} is the device noise figure defined at the selected generator impedance.

In these first three amplifiers, the indicated $\mu\beta$ considerations have resulted in first stage emitter resistances corresponding to Δ_{RS} of approximately 1 dB. Since the hybrid transformers used in these positions are equal ratio, the hybrid loss, allowing 0.5 dB dissipation loss in the transformer, is approximately 3.5 dB. For over-all NF of 10.5 dB,

$$NF_{trans} = 10.5 - 1.0 - 3.5 = 6.0 \text{ dB}.$$

Thus the first stage may be biased at as large a current as possible at which the device noise figure is 6.0 dB or less.

The "largest" current is suggested since, in a configuration of the type selected, nonlinear distortion originating in the input or low level stage will generally not be negligible. Consequently, it is desirable to bias the first stage so that, while satisfying noise figure objectives, nonlinear distortion effects are minimized. As is described in greater detail in Ref. 2 this can usually be achieved by increasing the dc bias current, which of course will generally degrade the noise figure performance.

It has been shown that the fourth amplifier of the A equalizer must provide 5 to 6 dB more gain than the others and that this is achieved by connecting a different input transformer of turns ratio 7:10+3. The ideal loss of such a transformer in the path to the transistor is 1.1 dB; the dissipative loss of the hybrid (primarily caused by core loss) is approximately 0.4 dB. Therefore, the total loss to the signal in the path to the transistor is about 1.5 dB. The increased closed loop gain required of this amplifier makes it possible to achieve satisfactory open loop transmission with no local feedback whatever on the first stage. Thus $\Delta_{RS} = 0$. Consequently the over-all amplifier noise figure will be 7.5 dB or less if the transistor noise figure is 6.0 dB or less, the same requirement imposed on the first stage of the other three amplifiers. Figure 11 shows the typical noise figure of this amplifier.

3.1.2.4 *Other Considerations.* The linearity of the feedback amplifier depends primarily on the inherent linearity of the transistors used and on the nature and amount of feedback applied. The effective transistor linearity depends largely on bias conditions and on connecting impedances. If the selected configuration causes the output or power stage to be the primary source of nonlinear distortion, advantage can sometimes be realized by using a common collector output stage. The degree of improvement (relative to a common emitter stage, for example) depends largely on the particular device characteristics and operating conditions; 5 to 10 dB advantages might be expected under the right conditions. Since a common collector output stage has already been shown to be desirable in the convenient realization of the adjustable equalizer networks, this connection for the output stage was used throughout the A (and the B) equalizer.

Negative feedback is used in the equalizer amplifiers to reduce nonlinear distortion and to reduce sensitivity to variations in the parameters of the active devices. It has already been pointed out that the configuration selected provides an isolation between adjustable equalizer networks which permits their totally independent operation and adjustment. The problems encountered in the realization of a satisfactory and stable $\mu\beta$ transmission in the equalizer amplifiers are very similar to those described in detail for the preamplifier of the basic repeater in Ref. 2 and are not elaborated here.

Also like the preamplifier, the equalizer amplifiers have a hybrid transformer connected at the input which largely determines the

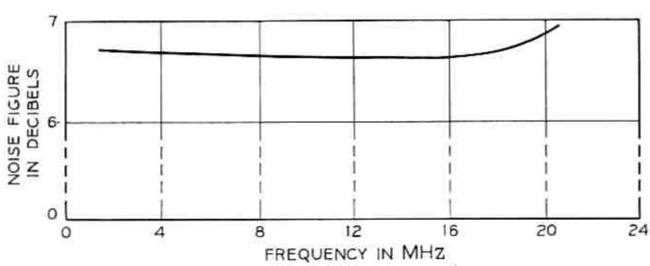

Fig. 11 — Typical noise figure of the equalizer amplifier A_4 of Fig. 6.

quality of the input impedance provided by the amplifier. The RLC network connected as shown in Fig. 8 is required to transform the "raw" impedance of the essential amplifier to that required for satisfactory and predictable performance. The capacitance is effective chiefly at low frequencies where it reduces the effect of the transformer mutual inductance, while the resistor-inductor pair provide a low Q compensation, effective primarily at high frequencies, for transformer inter- and intrawinding capacity.

3.1.3 *The Design of the Equalizer Networks*

3.1.3.1 *Bode Network Design.* The use of Bode networks (sometimes called Bode equalizers or Bode regulators) in transmission systems is not new.[7] One of the early applications was in the L-1 coaxial system.[8] The higher frequency range of the L-4 system introduces some unique problems resulting in design modifications. Before continuing, let us examine the basic relationships for Bode network design.

The three standard configurations using Bode networks are shown in Fig. 12. With R_0 as defined in Fig. 12 and with $Z_1Z_2 = R_0^2$, the insertion factor of all three configurations is

$$e^{\varphi} = 1 + \frac{Z_1}{R_0} = 1 + \frac{R_0}{Z_2}. \tag{7}$$

Normally the impedance Z_1 is realized as shown in Fig. 13(a); the impedance Z_2 is obtained as shown in Fig. 13(b). Equation (7) can be written

$$e^{\varphi} = 1 + \frac{1}{R_0(G_1 + Y_{1A})} \tag{8}$$

with

$$G_1 = 1/R_1, \qquad Y_{1A} = 1/Z_{1A}.$$

The corresponding insertion loss equation is

$$A = 20 \log_{10} | e^{\varphi} | = 20 \log_{10} \left| 1 + \frac{1}{R_0(G_1 + Y_{1A})} \right| \text{dB}. \tag{9}$$

By varying the termination shown in Fig. 13 between 0 and ∞, the loss A can be made to vary from $20 \log_{10} | 1 + R_1/R_0 |$ dB down to 0 dB with a characteristic shape determined by the shaping networks Z_A or Z_B. The variations will be symmetric if $R_0 = R_1R_A^2/(R_1^2 - R_A^2)$, resulting in a flat loss $A_0 = 10 \log_{10} | 1 + R_1/R_0 |$ dB.

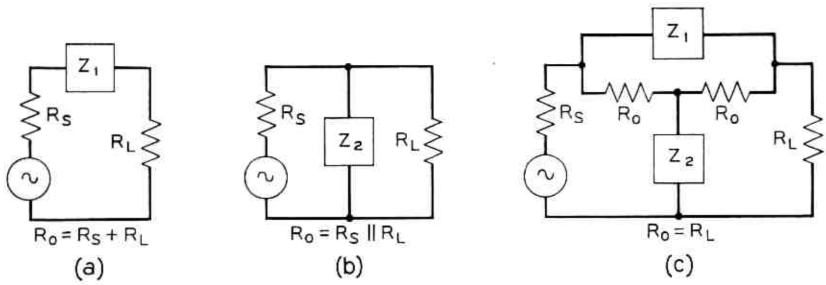

Fig. 12 — Basic equalizer network configurations: (a) series, (b) shunt, (c) bridged-T.

Equations (8) and (9) can be normalized so that the greatest loss variations are between -1 and $+1$ dB and the flat loss is 0 dB. Then charts can be used to relate G and B [where $Y_A = 1/Z_A$ (normalized) $= G + jB$] to the resulting normalized loss and phase. The network configuration N_A and the choice of G and B determine the ideal Bode network response.

The effect of the Bode network G parameter for the shape used in the A equalizer is evident from Fig. 14. A section with a G greater than 0.21 has no overshoot, but: (*i*) an undershoot remains at frequencies far removed from the desired response and (*ii*) the shape falls off too rapidly, resulting in ineffective equalization between the bump shape centers. (However, allowing a small loss where the response should ideally be zero permits the use of relatively high values of G). At low values of G the shape is reasonable between the crossovers, but a large overshoot can cause interference with adjacent shapes. The choice of G is thus a compromise. Values between 0.1 and 0.2 are used in the L-4 system equalizers.

In the design of L-4 A and B equalizer Bode networks, the bridged-T

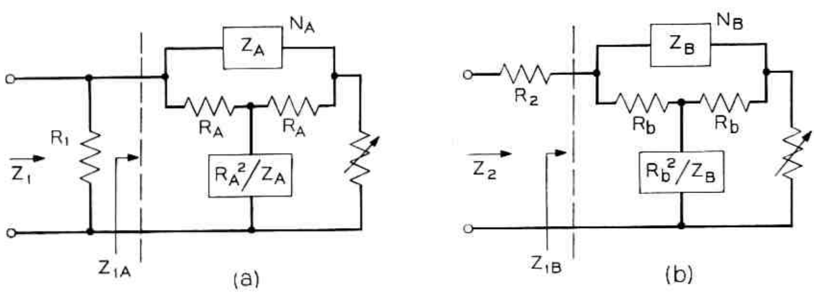

Fig. 13 — Realization of Z_1 and Z_2. (a) network Z_1 and (b) network Z_2.

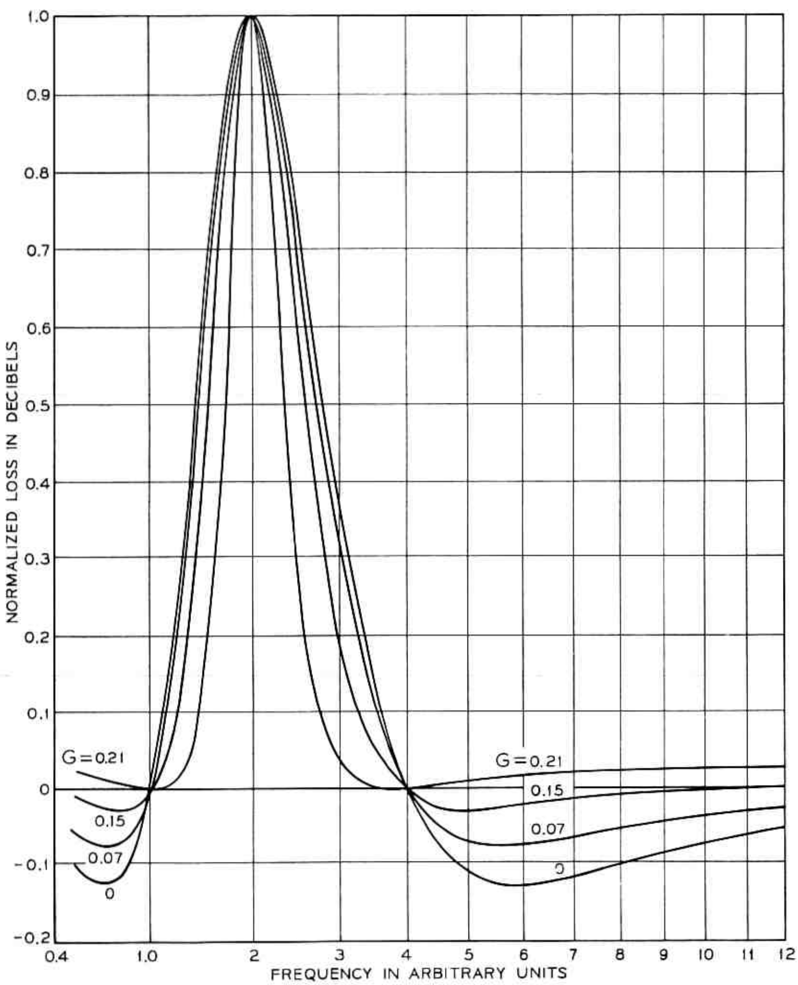

Fig. 14 — Effect of the Bode network parameter G on the response of the network.

configuration of Fig. 12(c) was eliminated from consideration by its complexity and by the undesirable necessity to provide two variable, tracking, inverse resistances terminating Z_1 and Z_2. The shunt arrangement was eliminated because the resulting impedance levels would require development of new thermistors or introduce major difficulties in amplifier design. Therefore, the series configuration (Fig. 15) is used for all of the A and B equalizer Bode networks.

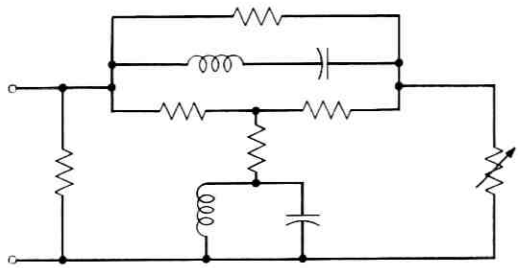

Fig. 15 — Ideal A equalizer Bode section.

For the A equalizer, the shaping network Z_A is a simple series resonance shunted by a resistor, resulting in the network Z_1 shown in Fig. 15. The variable resistance is supplied by a Western Electric 2A thermistor. The thermistor has a separate heater, substantially reducing signal coupling through the logic and memory circuitry. Thermistor characteristics at 140°F ambient temperatures vary as illustrated in Fig. 16. To maintain approximately equal loss changes in dB as the thermistor heater current is stepped by the memory, the linear portion of the characteristic is used as far as possible.

Realization of the Bode sections requires a deviation from the ideal parameters because of parasitics within the equalizer and impedance

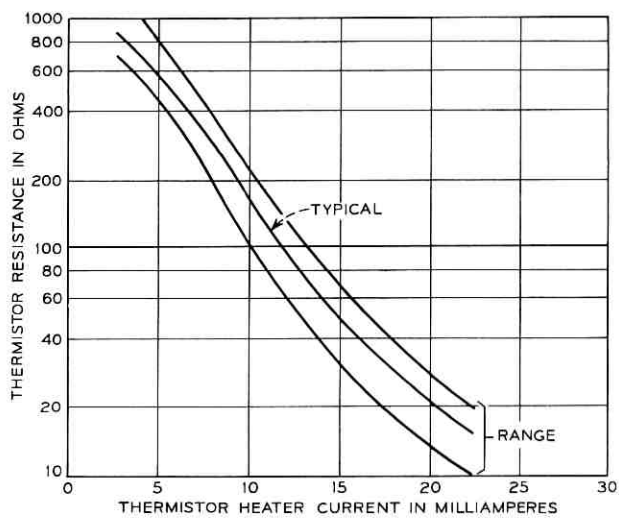

Fig. 16 — Thermistor R-I characteristics. Shown are the typical, maximum, and minimum characteristics.

interaction with the amplifier circuitry. The chief parasitics are the interwinding capacitance and inherent resistance of the inductors and the distributed capacitance and lead inductance associated with the thermistors. These undesired interactions produce a lowering of the realized G relative to the design G and introduce a distortion of the characteristic as shown in Fig. 17 for a feedback path Bode network. (The interamplifier Bode networks produce the same characteristic crossing shift, but the higher frequency loss response is decreasing rather than increasing.)

To overcome these difficulties, the component values were slightly altered based on laboratory measurements and a capacitance was placed in series with the thermistor leads (except for the highest and lowest frequency shapes, where the distortion is less of a problem). In addition, a capacitance was placed across the input of the Bode sections located in the amplifier feedback path.

3.1.3.2 *Trim Equalizer Design.* The amplifiers, transformers, and networks do not have a perfectly flat frequency response, and a fixed trim equalizer is required to provide an over-all flat nominal response for the A equalizer. The configuration for the A trim equalizer is shown in Fig. 18. The equalizer is composed of two bridged-T sections which provide loss peaks, one bridged-T section providing a loss valley, and a potentiometer for flat loss adjustment. The constant-resistance

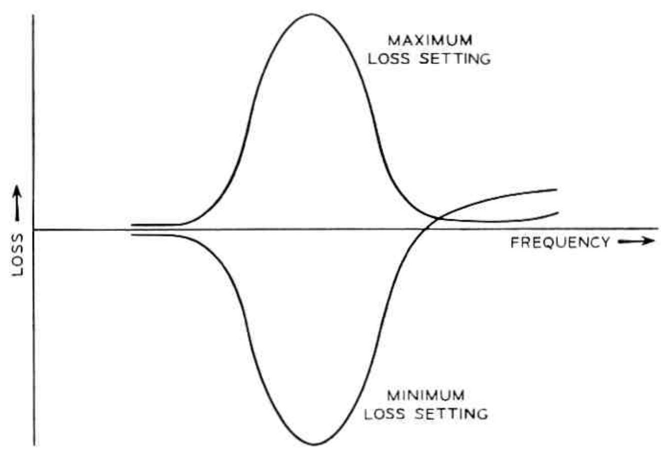

Fig. 17 — Distorted network response before compensation. The distortion results chiefly from circuit parasitics and has the effect of lowering the realized G relative to the design G.

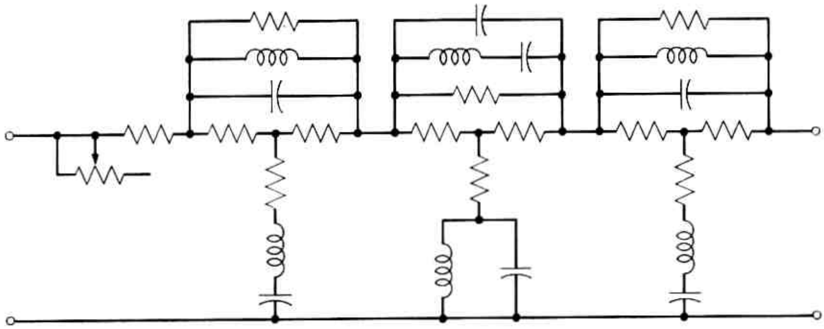

Fig. 18 — Configuration of the A trim equalizer.

bridged-T sections were designed with the assistance of specialized time-sharing computer routines and a digital computer general-purpose optimization program.[9]

The location of the trim equalizer between two amplifiers of the A equalizer renders its impedance relatively unimportant. Thus, the equalizer is not designed for high return loss. Because of the effect of parasitics, an extra capacitor is added across the bridge arm of the valley section. This capacitance resonates with the effective inductance of the series LC branch of the bridge arm, peaking the high-frequency side of the loss valley.

3.1.4 *Other Components of the* A *Equalizer*

The adjustable equalizer networks are controlled through the indirectly heated thermistor associated with each network. In order that these thermistors provide, for a given heater current, a predictable and constant termination to the network, it is necessary to establish a controlled environment. This is provided by placing the thermistors in an oven, the temperature of which is established and maintained by a proportional-control temperature-regulating circuit. The thermistors should be maintained at the lowest possible temperature consistent with operating environment, since this permits a maximum possible variation in thermistor resistance for a given variation in heater current. Consequently, the oven is maintained at approximately 140°F and provides a virtually unchanging environment for the thermistors between about 60°F and 140°F ambient temperatures. (The typical resistance heater current characteristic for the thermistors at 140°F was shown in Fig. 16.)

The input and output connecting circuits of the A equalizer permit the insertion of the test signals which are used at the remote control center to determine either the gain settings of the equalizer networks or the misalignment of the section of the system between equalizing stations. These, in Fig. 6, are designated H1-Net 1 and H2-Net 2, respectively. The loss-frequency characteristics of the two circuits are carefully specified and controlled so that the test signals originating in the test oscillator circuit are applied to the L-4 line at a transmission level of −20 dBm0. The input and output connecting circuits are thus designed so that a comparison of the test signal level measured at the equalizer output (or any point beyond, such as at the control center) when the test signals are applied first to Net 1 and then to Net 2 provides an accurate measure of the A equalizer gain at the test signal frequencies (denoted by the arrows on Fig. 5).

Correspondingly, a comparison at the control center of the received test signal level when applied first to Net 2 of an A equalizer and secondly to Net 1 of the A equalizer next closer to the control center provides an accurate measure of the response at the test signal frequencies of that section of the system between the equalizers. This will usually be a section including about 25 basic and regulating repeaters. The information gathered in this fashion is then used to determine the optimum settings of the several equalizer networks.[1]

3.2 *Equalizer Remote Control Circuits*

Equalizer settings are remotely controlled by the transmission from the L-4 control center of coded commands which at the equalizer location must be detected, decoded, and acted upon. These functions in the equalizing repeater involve: (*i*) the command receiver, which selects those commands from the line signal which are transmitted in the command channel assigned to that location, and which recovers the originally encoded audio signals; (*ii*) the logic circuit, which performs an analog-to-digital conversion of the command-receiver output and decodes the commands inherent in the signals received; (*iii*) the test oscillator circuits which provide the test signals used to measure the gain of the A equalizer and the misalignment of the cable sections between equalizers; (*iv*) the memory circuits which maintain the settings of the equalizer networks between adjustments; and (*v*) the monitoring oscillator power supply and switching circuit which provides, on command, the 15 mA dc required to energize a string of monitoring oscillators, and which connects this dc source to one of the two oscillator strings adjacent to the equalizing repeater.

The command receiver is shown in block form in Fig. 19. The input

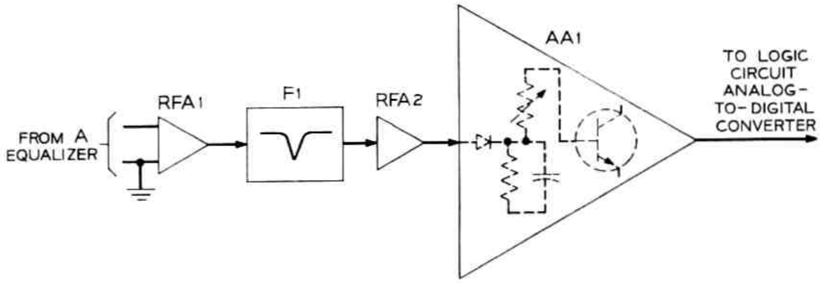

Fig. 19 — Block diagram of the command receiver. The amplifier AA1 includes the envelope detector and amplitude control.

signal to the receiver is bridged from the output of the first amplifier of the A equalizer. The combination of amplifiers and filter permit the receiver to be adjusted so that, with nominal input to the A equalizer, approximately 10V p-p is delivered to the logic circuit.

The logic circuit consists of an analog-to-digital converter and a series of gates and drivers. The analog-to-digital converter (Fig. 20) includes ten vibrating reed selectors and ten Schmitt trigger circuits. Each of the reed selectors has a resonant frequency which corresponds to one of the audio frequencies used by the control center to generate the various remote control system commands.

When the series-parallel connection of the selector windings, which forms the command receiver load, receives a signal of adequate amplitude at one of the these frequencies, the corresponding selector responds with intermittent closure of a pair of contacts at a rate equal to its resonant frequency.[10] The vibrating contact is connected in series with an RC charging circuit which in turn is connected in the base of the Schmitt trigger circuit. The presence of the particular audio frequency is thus detected by the reed selector and causes the operation of the trigger circuit which remains operated for the duration of the audio signal. Most of the control system commands are transmitted in bursts of approximately 300 ms duration and these commands correspondingly produce, at the analog-to-digital outputs, pulses of approximately the same length. (The only commands not falling into this category are those associated with test or monitoring oscillator turn-on which are continuous during the interval that these oscillators are "on.")

The outputs of the analog-to-digital converter associated with equalizer adjustment are connected to the several gates (Fig. 20).

Each gate is associated with a particular equalizer-adjust command and is connected to one of the six memory circuits. The equalizer-adjust commands advance (count-up) the count of the memory; retard (count-down) the count of the memory; or cause all of the memories to go to the 011111 state (reset).

Analog-to-digital converter outputs G and H are connected to driver circuits which in turn operate the relays K1 and K2, respectively, of Fig. 21. The contacts of these relays apply dc power to the test oscil-

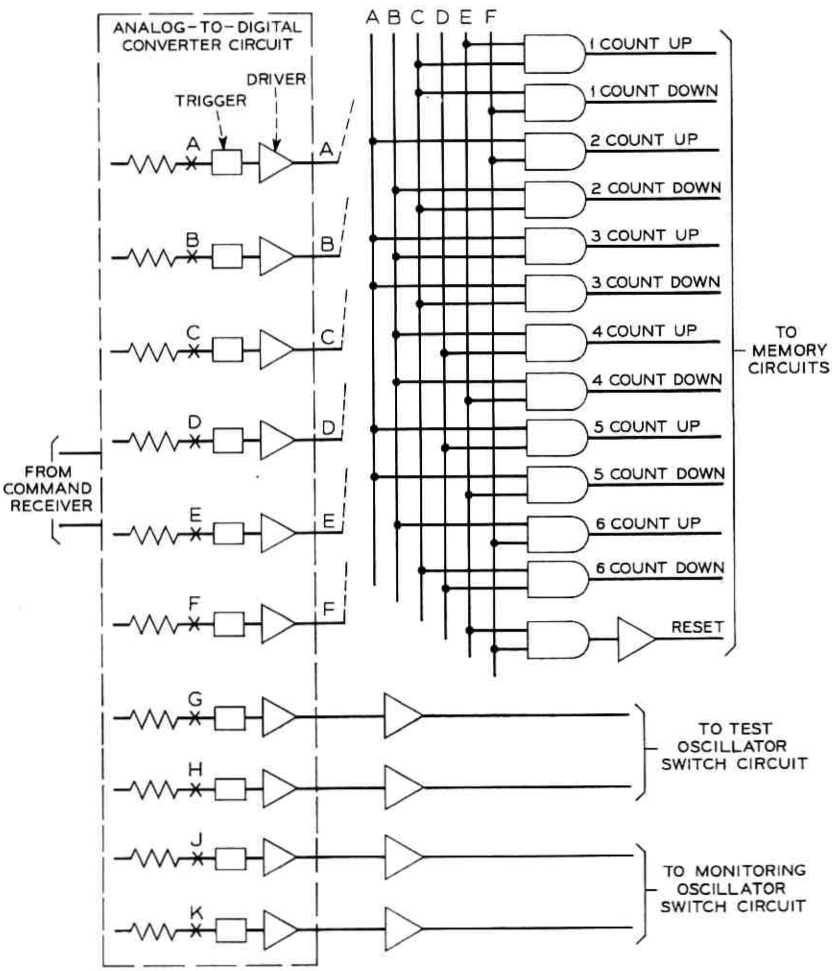

Fig. 20 — Block schematic of the A logic circuits.

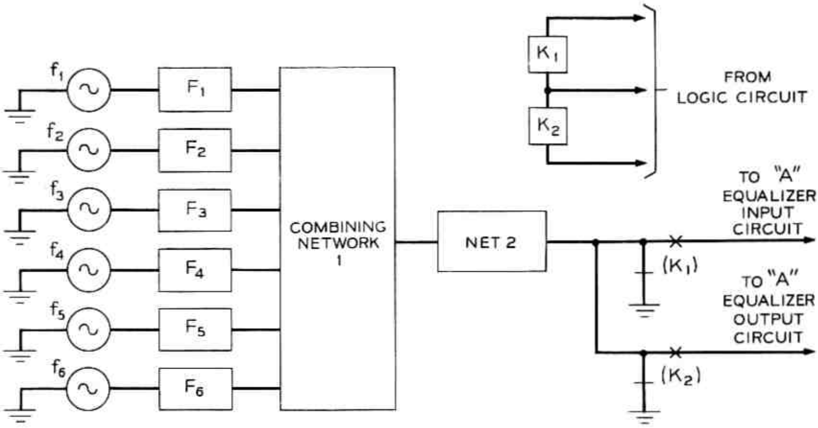

Fig. 21 — The A test oscillator circuit.

lators and connect the test oscillator signals either to the input (Net 1 of Fig. 6) or to the output (Net 2 of Fig. 6) of the A equalizer.

Outputs J and K are connected to circuits which, in a similar way, cause power to be applied to one or the other of the two adjacent groups of monitoring oscillators when the appropriate command is received.

The test oscillator circuit is shown in Fig. 21 and provides at a carefully controlled amplitude the six test signals which ultimately provide the basis for the settings of the six A equalizer networks. The frequencies of the test signals span the L-4 band and are indicated in Fig. 5 by the arrows associated with shapes 1 to 6. As can be seen, each is at or near the center-frequency of the associated "bump." The outputs of the six oscillators are individually filtered (F1 to F6) before the six signals are combined (combining network 1) and the proper level is established (Net 2). The combined signal is connected to the switching circuit consisting of K1 and K2 which, upon command, will apply power to the oscillators and connect the test signals either to the input or to the output of the A equalizer.

Associated with each adjustable equalizer network is a memory circuit which maintains the equalizer settings between adjustments. This is done by using the memory to establish and maintain the current supplied to the thermistor heater in the corresponding network. The L-4 equalizer memory is indicated in Fig. 22 and consists of a six-stage reversible binary counter which can established 64 different

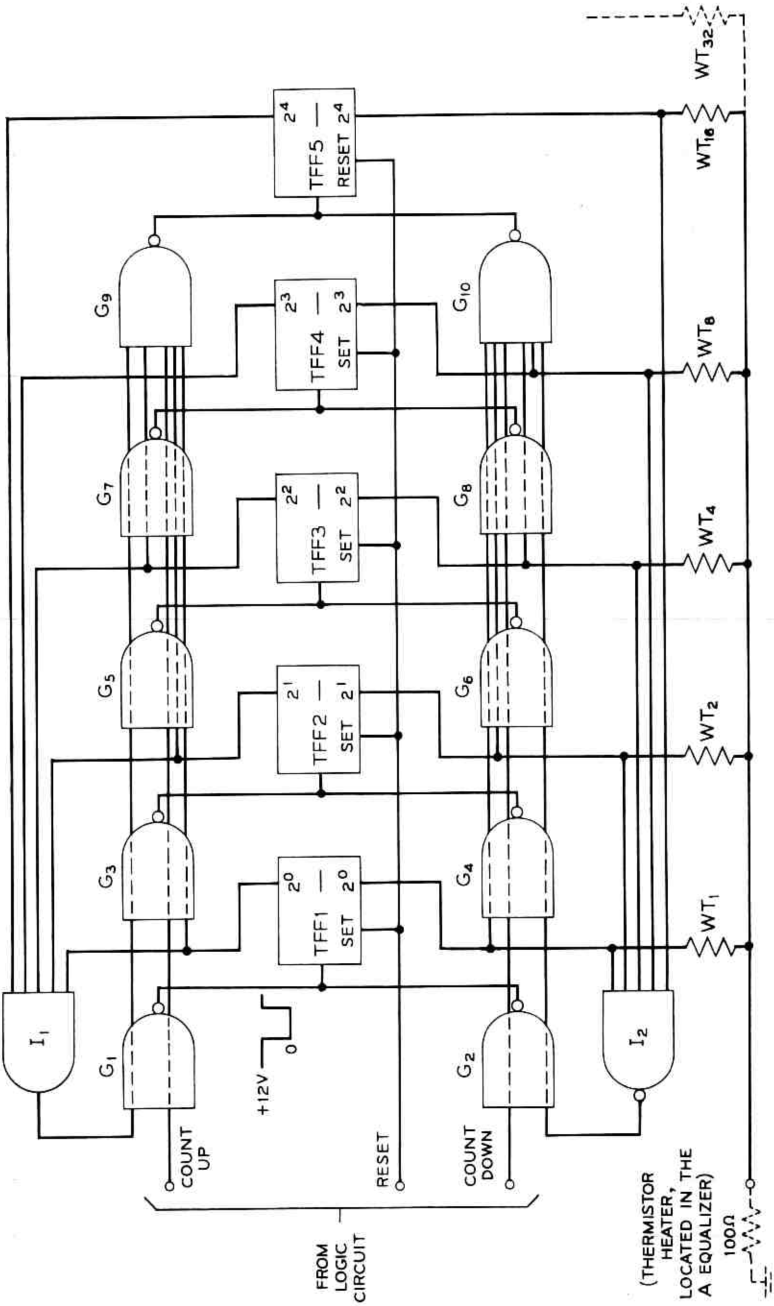

Fig. 22 — Memory circuit for the equalizers. This circuit is used throughout the system for the control of A and B equalizer settings.

values of heater current and consequently 64 different settings of equalizer network gain.

As the memory is counted by remote commands originating in the control center from the 000000 state to the 111111 state, the current supplied to the network thermister increases from minimum to maximum. Depending on whether the network involved is connected in series with the amplifiers of the A equalizer or connected in the β-circuit of the same amplifiers, the gain of the equalizer goes to maximum or minimum, respectively. In the simple or "uninhibited" binary counter one pulse beyond the 111111 state lies the 000000 state (and vice versa); such a change in this application would cause the corresponding equalizer network to go from one extreme gain setting to the other (which may be as much as 12 dB different at peak frequency). Consequently, over- and undercount inhibit are designed into the counter. Having counted up to the 111111 state or down to the 000000 state, further attempts to count in the same direction will produce no change in either the counter or the associated network. This is accomplished (Fig. 22) by the gates I1 and I2 which shut off the inputs to the counter associated with counting up and counting down, respectively.

The several equalizer bumps have adjustable range of approximately ± 4 or ± 6 dB. Thus the difference in the gain associated with consecutive states of the memory average about 0.1 to 0.2 dB. This places an ultimate limit on how close to ideal the response at the several test frequencies can be made.

The dc-dc converter used to power the monitoring oscillators controlled by a particular station is described in Ref. 11. The switching circuit associated with fault location is shown in Fig. 23. When the proper command has been received from the control center and decoded in the logic circuits, the 0.5A dc line current is switched to the dc-dc converter. The resultant 15 mA dc from the converter (at up to 260V) is connected to the group of monitoring oscillators either to the east or to the west of the equalizing station (depending on which command had been received). Each of these groups may include oscillators at up to fifteen repeater stations in addition to the equalizing station. A typical main section is divided into six such groups, two groups controlled by each of the two equalizing repeaters and one group controlled by each main repeater. The converter operates only when fault location is under way, and in the absence of one of the required commands the dc-dc converter input is short-circuited by the oscillator switch circuit as shown in Fig. 23.

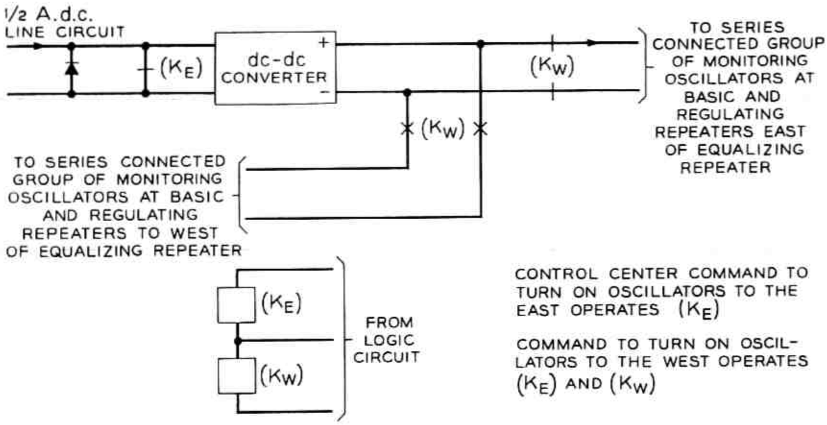

Fig. 23 — Monitoring oscillator power and switching circuit.

3.3 *Physical Design*

The entire L-4 equalizing repeater complex occupies the volume of two apparatus cases of the same size and shape used for the basic and regulating repeaters.[12] For a fully equipped Coax 20, there are 40 of these sealed pressurized apparatus cases which house the transmission equipment required at an equalizing station.

Because of the repetition in circuitry of the L-4 repeaters, the basic building block concept was carried to the physical design of the equalizing repeater. The same elements present in the regulating repeater appear again in the equalizing repeater with the addition of the A equalizer and the equalizer and fault-location control circuitry. Thus, the regulating portion of the equalizing repeater consists of an L-4 regulating repeater modified to distribute power to the other plug-in units.

The remaining parts of the equalizing repeater are separated into two types of equipment packages—large castings with over-all dimensions about 3.4 by 12 by 17 inches which occupy the central portion of the apparatus cases, and smaller, peripheral units filling spaces of about 2.5 by 6 by 15.5 inches along the sides of the larger units (see Fig. 24). Because of the series powering scheme for the L-4 line, all the circuits of the equalizing repeater are at high voltage with respect to sheath or earth ground. High voltage points have been made as inaccessible as possible and all electronic components, active and passive, are contained within epoxy-insulated cavities of the plug-in units.

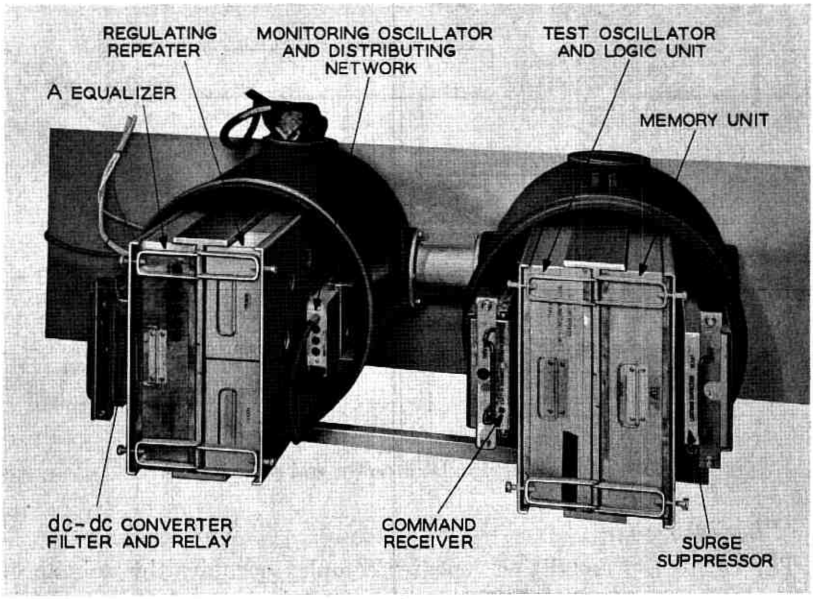

Fig. 24 — Equalizing repeater apparatus case. The left cabinet houses the transmission circuits; the right cabinet houses the remote control circuits.

The A equalizer assembly (Fig. 25) contains the equalizer amplifiers, Bode networks, thermistor ovens, temperature control networks and the regulating diodes which provide the several dc voltages required by these circuits and the control circuits. Because the lead lengths interconnecting the amplifiers and Bode networks are critical, the networks have been made an integral part of the amplifier assembly which uses an epoxy glass printed wiring board mounted in a cast aluminum housing. These are the characteristic die cast amplifier housings used throughout the L-4 designs where the bottom amplifier cover, containing the mounting studs for the amplifier, is bonded to the epoxy coated cavity in the main casting using an epoxy sheet adhesive.

The same bonding technique is used in the test oscillator and logic assembly to mount the six A test tone oscillators (see Fig. 26). The individual oscillator containers are bonded into position as close as possible to their respective filters. The array of cast ribs in the cavity of the main housing shields the small filters.

The logic circuitry, mounted in the test oscillator and logic assembly, uses a series of reed selectors which respond to the audio command tones from the main station control center. Since the conventional mounting arrangements of these vibrating reed selectors in sockets could not be adapted to the L-4 designs, the selectors were soldered on epoxy-glass printed wiring boards. Clamping arrangements and a sequential assembly method were developed to prevent relative motion between the selector's permalloy shell and its molded plastic base, and between its base and its terminal leads to the energizing winding. These expedients permit normal handling and manipulation of the selectors during shop assembly, testing and inspection.

The memory assembly houses six memory circuits on epoxy glass printed wiring circuit boards. The memories are mounted on standoffs in individual cavities formed in the main casting.

The command receiver (see Fig. 27) typifies the type of construction used in the peripheral units. They have fabricated housings, insulated with an epoxy coating; constituents are mounted by standoffs and by bonding. They are a single cavity deep, lighter than the larger cast type units, and have guide rails on their sides for mating with guides on the apparatus case chassis. Handle assemblies contain locking features for securing the units in place. During hardness

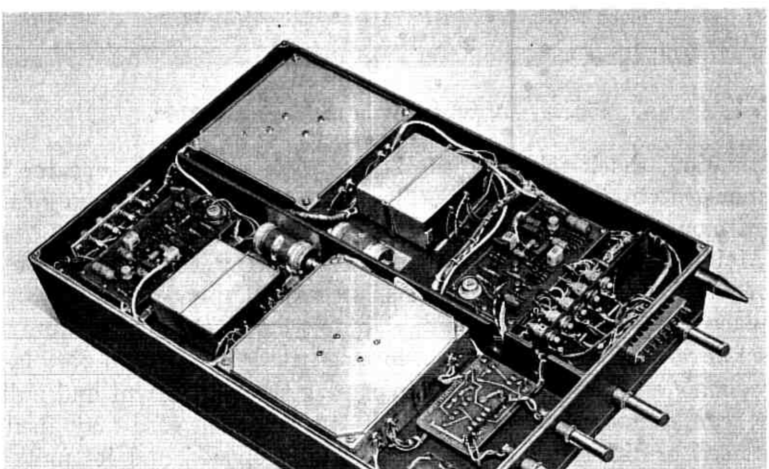

Fig. 25 — A equalizer assembly.

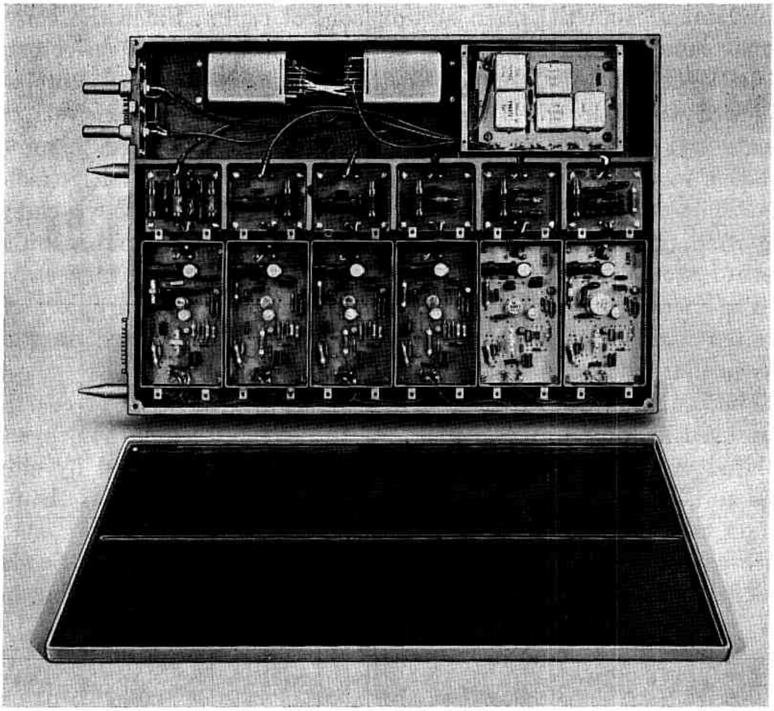

Fig. 26 — A test oscillator and logic assembly.

evaluation, shock tolerance of the complete equalizing repeater was demonstrated at the 50g level. Therefore, this equipment qualifies for hard-mounting in manholes.

Maintenance of equalizing repeaters is very similar to that for basic and regulating repeaters. If necessary, plug-in units can be replaced under power without risk to personnel. A procedure has been developed to enter the transmission case first, and to disable the equalizing repeater complex by bypassing the line current and pulling the regulating repeater. Units in the transmission case may then be replaced or the control case may then be opened and its units replaced. Auxiliary aids or tools are available which permit patching, removing pads, discharging high voltage capacitors within the repeater and A equalizer, and monitoring the performance of the repeater. If maintenance of the apparatus case is necessary, the line power is turned down so that the apparatus case chassis may be removed safely.

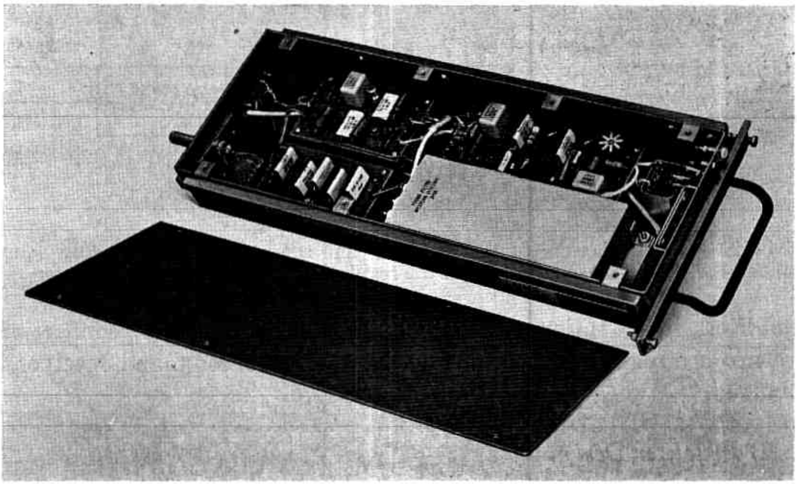

Fig. 27 — Command receiver assembly.

IV. MAIN STATION REPEATERS

The main station repeaters are made up of the several major elements indicated in Figs. 2 and 3. The line transmitting repeater (Fig. 28) and the line receiving repeater (Fig. 29) taken together perform the service of a single regulating repeater.[2]

The A equalizers of the main repeater are identical to those of the equalizing repeater in all essential respects, differing only in physical design and manner of powering. The remote control aspects of the main repeater differ from the equalizing repeater only in phyiscal design and in the absence of a command receiver. As described in Ref. 1, the command receiver for the main repeater equalizers (both A and B) is a physical part of the command looping panel. In receiving

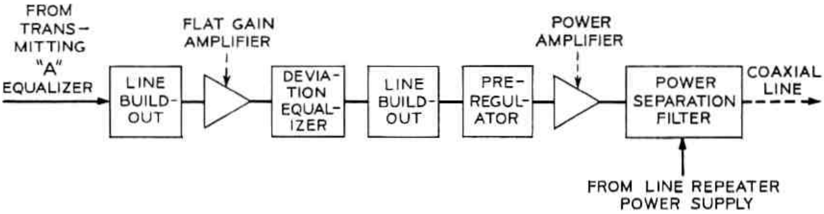

Fig. 28 — Block diagram of the transmitting line repeater. This is roughly equivalent to the preregulating portion of the regulating repeater.

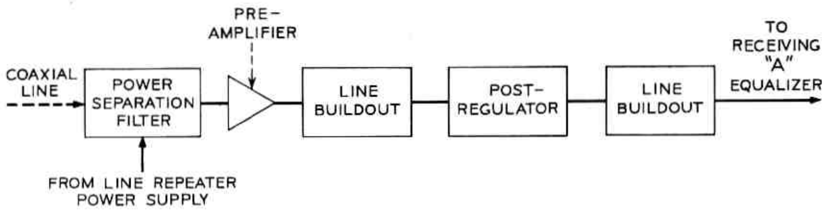

Fig. 29 — Block diagram of the receiving line repeater. This is roughly equivalent to the postregulating portion of the regulating repeater.

equalizers in a station which includes a control center, connections are made directly to the control center and no command receiver is involved.

4.1 B *Equalizer Design*

The design of the B equalizer and its associated control circuits in most ways closely parallels that of the A equalizer. This section concentrates on the areas of significant difference: (*i*) the equalizer network design and (*ii*) the manner in which the test signals used for B equalizer adjustment are applied. Figure 30 is a block diagram of the B equalizer.

4.1.1 *Design of the Equalizer Networks*

The B equalizer requirements differ somewhat from those of the A equalizer:

(*i*) The minimum desired loss range is ± 3 dB.

(*ii*) Instead of six broad shapes, the B equalizer consists of ten narrow shapes.

(*iii*) To keep amplifiers at a minimum, it is desirable to replace the two-section-per-amplifier scheme used in the A equalizer with an arrangement allowing more sections per amplifier.

The first requirement eases the design of the B equalizer; the second and third requirements introduce new problems.

The four shapes near the center of the L-4 band are provided by Bode networks in the feedback paths of the amplifiers in the manner of the A equalizer. Each of the four networks includes a capacitor in series with the thermistor leads and another capacitor shunting the input to the network.

To conserve on amplifiers, the single bump series Bode network placed between the amplifiers of the A equalizer is replaced with a

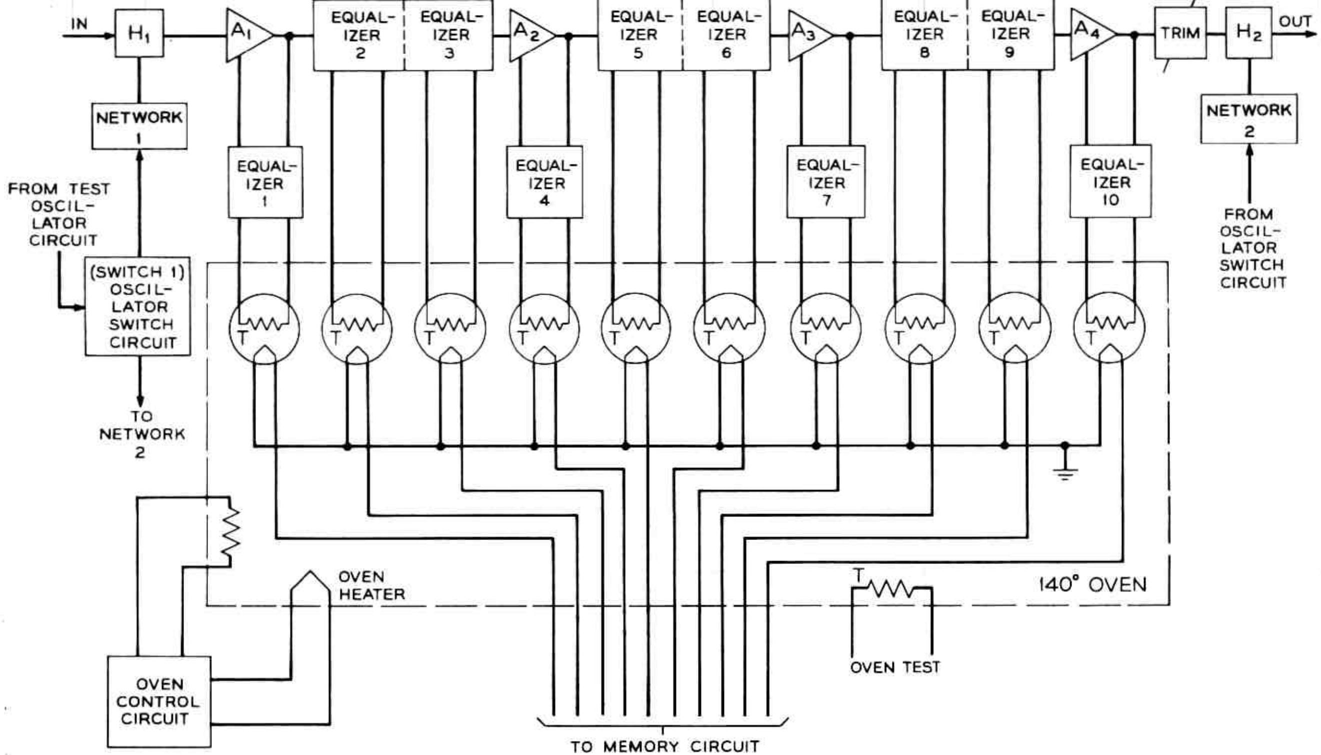

Fig. 30 — Block schematic of the B equalizer. For clarity only one oven and control circuit is shown. The B equalizer circuits are actually distributed among four plug-in modules and the rack-mounted shelf. Each module has its own oven and control circuit for the thermistors required by the equalizing networks in the module.

double-bump series configuration of the type described by Lundry.[8] As is illustrated in Fig. 31, the network consists of a frequency separation network and two Bode networks. One Bode network provides a shape near the low edge of the L-4 frequency spectrum, while the other network supplies a high frequency shape.

Resistor R_1 corresponds to the same resistor in Fig. 13(a). The frequency selective network is designed so that one branch is a low impedance and the other a high impedance at and above the effective frequency band of one Bode network as shown in Fig. 31(c). Between the effective ranges of the Bode networks, the impedances of the frequency selective networks are rapidly varying. These networks are inverse about R_2, which is also the nominal impedance at the input

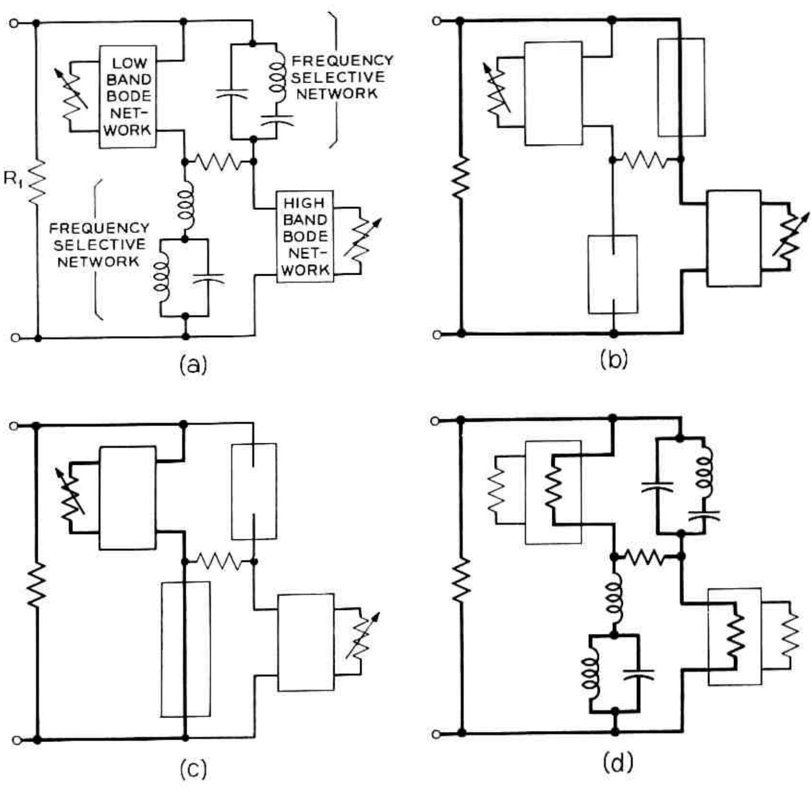

Fig. 31 — (a) Double-bump Bode network; (b) high frequency equivalent; (c) low frequency equivalent; (d) midband equivalent.

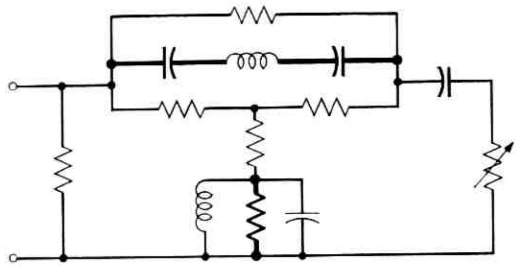

Fig. 32 — High frequency B equalizer Bode network.

of the Bode networks. Thus, in mid-range, the network is a properly terminated, constant-R, bridged-T section as shown in Fig. 31(d).

The high frequency networks, however, are modified in three ways as shown in Fig. 32. The first modification is to add a capacitor in series with the thermistor lead to remove the tilt and crossing variation in the characteristic. The second modification is to replace the capacitor in the bridge arm with two capacitors because the desired capacitance is small (4.65 to 12.45 pF). The third change consists of adding a resistor in parallel with the tank circuit in the shunt arm. The resistance serves a Q-balancing function. Above the center frequency of the bump shape, the tank circuit capacitor and the bridge network inductor predominate, but the capacitor has a considerably higher Q. Without the Q-balancing resistor, the Bode network nominally-flat shape is distorted as shown in Fig. 33. This distortion is

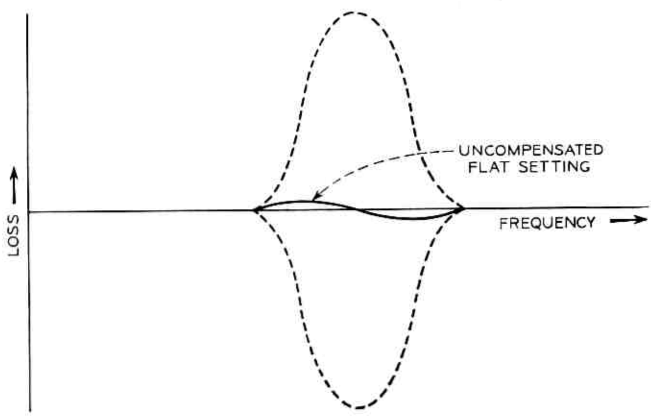

Fig. 33 — Effect of Q imbalance.

present in all of the A and B equalizer shapes, but is negligibly small except for the high-frequency B shapes where the uncompensated peak-to-peak distortion is about 0.3 dB.

4.1.2 *Trim Equalizer Design*

Just as a trim equalizer is needed to remove residual fixed deviations in the A equalizer, a similar equalizer is provided to flatten the response of the B equalizer. Both trim equalizers include one loss valley and two loss peak shaping sections and one level adjustment section. The configuration for the B trim equalizer is shown in Fig. 34. One resistor has been saved in each bridged-T section by performing a wye-T transformation on the bridge resistor and the T resistors and then combining the remaining two resistors in the shunt arm.

Because the trim equalizer is located at the output of the B equalizer, its return loss is important. Hence the simple potentiometer level adjustment of the A trim equalizer is replaced by a T-pad including a potentiometer.

4.1.3 B *Equalizer Test Signal Oscillators*

Unlike the A equalizers, all of the B equalizers in a particular main repeater setup—for a coax 20 this means up to ten receiving and ten transmitting B equalizers—share a single set of test signal oscillators which, on command from a control center, is connected to the B equalizer of the line being equalized. Since the transmitting and receiving equalizers are controlled by different control centers, and since each of these control centers should be able to equalize independently of the other, the test signals can be connected simultaneously to one transmitting and one receiving B equalizer.

The B test oscillator and oscillator-connecting circuits are shown schematically in Fig. 35. The control center command is directed to

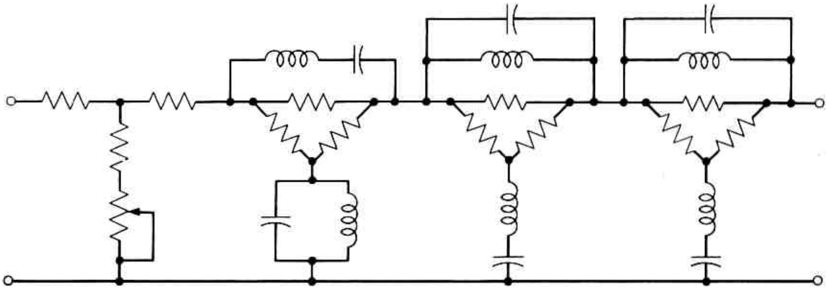

Fig. 34 — B trim equalizer.

the logic circuit of the B equalizer to be adjusted where the command is decoded. In the case of an oscillator turn-on command, the logic circuit establishes the dc voltage required to operate the oscillator switching circuit so long as that command continues to be received from the control center. The switch circuit (switch 1 of Fig. 30) establishes a path to either the input or the output of the B equalizer, depending on the command received, and operates relay K_1 or K_2 (Fig. 35) in the B oscillator and connecting circuit, depending on whether a transmitting or receiving B equalizer, respectively, is involved. (Not evident in Fig. 35 is the fact that operation of either K_1 or K_2 also applies dc power to the set of oscillators, the oscillators being normally in the OFF condition.)

Like the A test signals, the B test signals are applied to the system at -20 dBm0. This is accomplished by the careful control of the oscillator amplitudes and of the insertion loss of the input and output connecting networks (networks 1 and 2 of Fig. 30) located at the B equalizer.

4.2 *Band-Edge Regulator*

The role of the band-edge regulator in the L-4 system is the dynamic equalization of the lower edge of the spectrum during the interval between A and B equalizer adjustments. The temperature dependence of the system response in this part of the transmission band is appproximately three times greater than in any other part of the band. In the absence of the regulator, the equalizer adjustment interval would be wholly determined by these low end effects and would necessarily be shorter.

The design of the regulator loop follows very closely the design described in detail in Ref. 2 for the cable temperature regulator of the line regulating repeater. A block diagram of the regulator circuit is shown in Fig. 36.

4.3 *Monitoring Tone Blocking Circuit*

The monitoring tone blocking circuit is connected between the band-edge regulator and the automatic line switch in the receiving main station repeater bay only (Fig. 3).

This circuit provides for (*i*) flat gain required to establish proper system levels, (*ii*) the fixed equalization of the coaxial cable running from the receiving bay to the automatic line switch (located in the control connecting bay) and back, and (*iii*) blocking the monitoring tones used in fault location when this function is required.

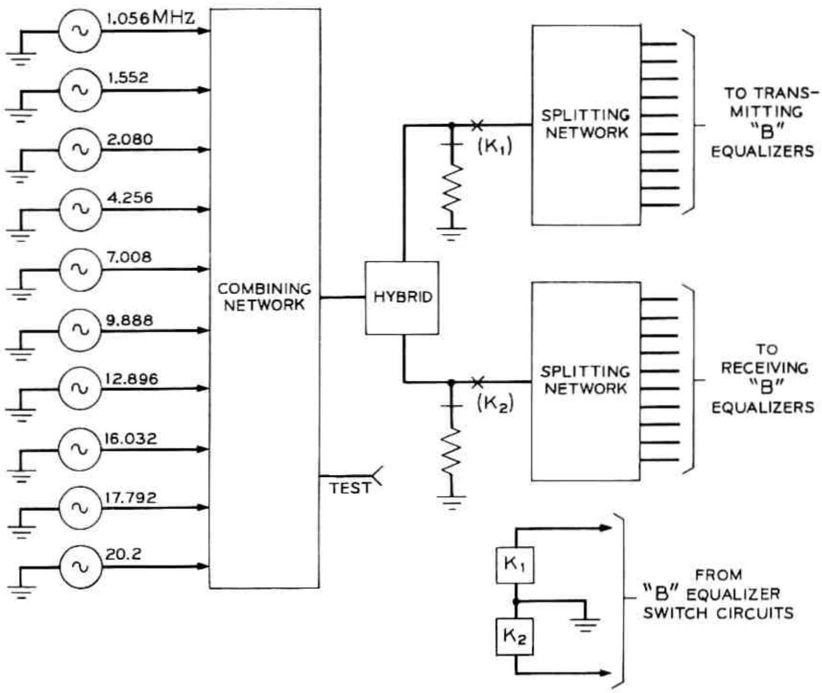

Fig. 35 — B test oscillator and oscillator connecting circuit.

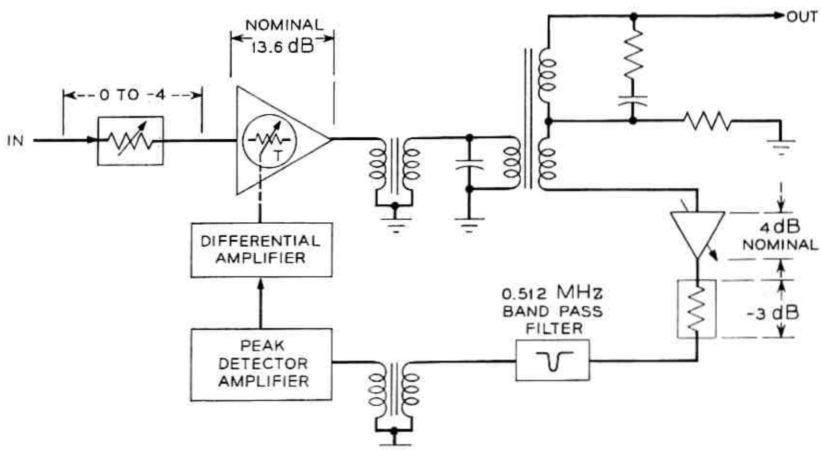

Fig. 36 — Block diagram of the band-edge regulator.

Figure 37 is a block diagram of the monitoring tone blocking circuit. The blocking filter option is exercised at those main stations having a control center, while the pad is used at power-feed stations. In the power-feed stations this permits fault-locating signals originating in the far section of a two-section control link to pass through the power-feed station and then to be observed at the control center at the receiving end of the near section.

The cable equalizer is selected from among a family of such equalizers which are available to compensate for the slope of the interconnecting cable losses. The selection depends on the length of cable to be equalized and is made to the nearest 25-foot multiple.

The nominal 15 dB flat gain amplifier is of the type used throughout the line connecting circuit and mastergroup multiplex. The gain of the amplifier is adjusted so that 12.5 dB of flat gain is provided between the output of the B equalizer and the block signal output jack of the line connecting circuit.[13] The gain of this circuit supplements the nominal gain of the band-edge regulator in achieving this goal.

4.4 *Physical Design*

Main station repeaters are the terminal elements of the L-4 repeatered line and are located in central offices. For hardened L-4 routes, these offices are underground buildings fully hardened to survive the designated overpressure.

Main station repeaters are designed for relay rack-type mounting on standard 23-inch, unequal flange, duct-type bays (11 feet, 6 inches high for most applications; 10 feet, 6 inches, and 9 feet high for offices with limited ceiling height). Two transmitting main station repeaters

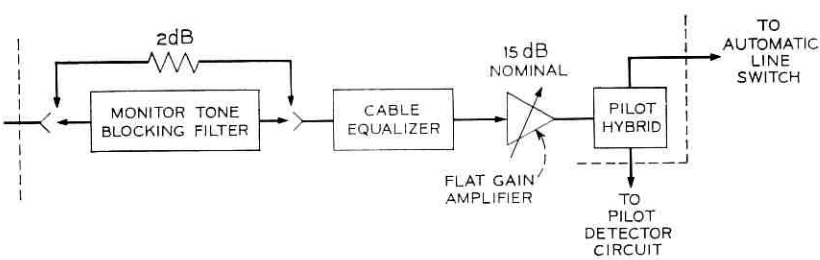

Fig. 37 — Block diagram of the monitoring tone blocking circuit. One of the pilot hybrid outputs is connected to the receiving line switch while the other goes to the switch initiating circuits.

are located in one transmitting bay while a single receiving main station repeater occupies a receiving bay.

Because the circuitry used in the line repeaters is required again in the main station repeaters, the concept of repetition in packaging arrangements was carried on to the physical design of the L-4 main station repeaters. Thus there are two distinct types of physical designs featured within the line bays—manhole-type packages, adapted for use on relay racks, and panel-type packages, designed specifically for relay rack mounting.

The line bays were styled to appear flush from the front. Panels and shelves are assembled on the rear small flanges of the 5-inch deep bay uprights and reach forward to match the closed duct formed in a bay lineup by the larger front flanges of the bay uprights. The equipment extends 10 inches to the rear of the bay framework for an overall depth of 15 inches excluding guard rails (see Fig. 38). Aisle space must be provided for access at the rear of the bays, making L-4 line bays unsuitable for back-to-back mounting. Equipment bay lineups are generated in multiples of four coaxials starting with the first receiving bay (for the first receiving coaxial) then a transmitting bay (for the first and second transmitting coaxial) and another receiving bay (for the second receiving coaxial). Equipment common to all of the coaxial lines in the cable is mounted in the control connecting bay, appearing typically in the middle of an L-4 equipment lineup. Thus, one Coax 20 requires a sixteen-bay equipment lineup comprising fifteen repeater bays and one control connecting bay. Provisions are

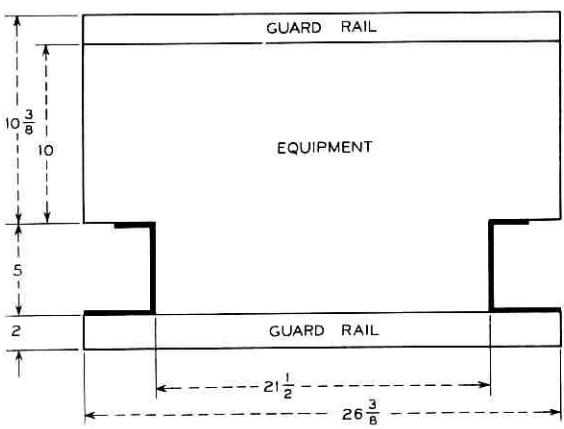

Fig. 38 — Plan view of a line bay.

made for shock isolating the bays by suspending them on shock mounts when conditions require. There was no special attempt to make the bay equipment rugged beyond a minimum 3g acceleration shock tolerance normally required for telephone equipment.

Typically the bays mount equipment in fixed panels, sliding shelves and panels with sliding drawers depending on the degree of access required for installation or maintenance. Wherever possible, the panels and shelves plug in to the local cable in the bay ducts.

The panels and shelves use light construction extensively. They are fabricated from 0.060 inch aluminum sheet and welded together using a generous number of ribs, struts, and stiffeners. The end results are bays equipped with sturdy, lightweight, compliant assemblies aesthetically pleasing because of structural simplicity.

4.4.1 *Transmitting Main Station Repeater*

Figure 39 shows the line transmitting bay equipment arrangement. The power separation filter (one per coaxial) at the top of the bay is shelf-mounted (the Fig. 40) and contains the filters and blocking capacitor needed to combine power and signal for transmission over the line without adversely affecting the line transmitting repeater. Access to the line is by solid dielectric coaxial connection to the cable terminal located in front of the bay lineup. RG 213/U cable, in rigid conduit or rigid raceway, is used for the power run from the high voltage dc-dc converter in the power room. Flexible conduit covers the power cable on the short run from the rigid raceway to the power separation filter where the power cable is hard wired by the installer. Solid dielectric insulation, metal shield plus outer insulation for the power cable, and the run in conduit are safety features to protect personnel and equipment. The power separation filter is so designed that inadvertent access to dangerous voltages is virtually impossible.

The power separation filter at the top of the bay is the only place where high voltage appears in the bay. For personnel safety, the equipment in the bay is not powered from the high voltage line but uses dc-dc converters powered from −24V battery to supply a quiet, regulated −25V source for distribution over the bay. Four converters are plugged into the rear of the fuse panel (see Fig. 41). Fuses are arranged in fuse blocks with four separate buses to provide A- and B-battery power to the equipment for the two coaxial cables in the bay. Equipment for the odd transmitting coaxial cable is powered from the two A buses and the equipment for the even transmitting coaxial cable is powered from the two B buses. Decentralized filter capacitors

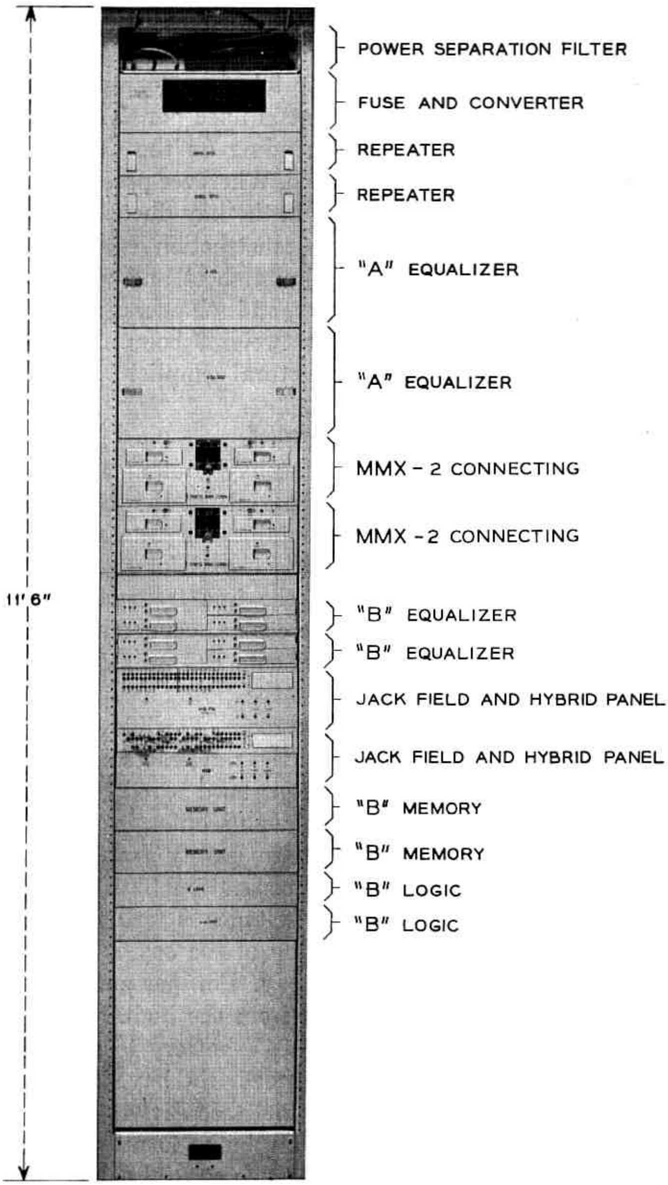

Fig. 39 — Transmitting bay equipment arrangement.

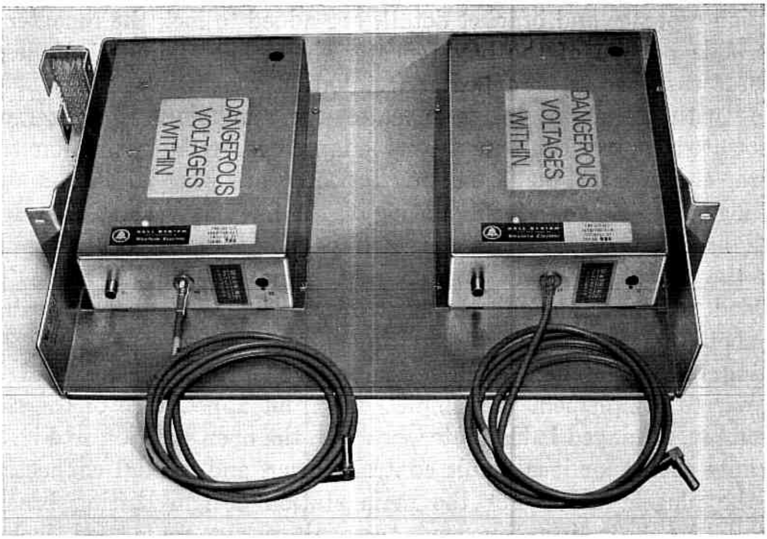

Fig. 40 — Transmitting bay—power separation filter shelf.

and the fuse alarm relay printed circuit board are also mounted in the fuse panel. The decentralized filter coil is mounted on top of the bay framework on standard mounting plate details. The dc distribution from −24V battery is made from the main power board in the office directly to the line bays.

The line transmitting repeater shelves (one per coaxial) are below the fuse panel to minimize the lead lengths to the power separation filter. The manhole regulating repeater casting is used for the main

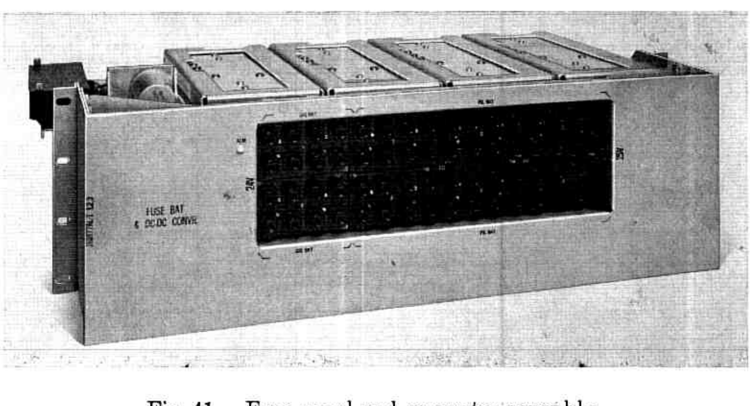

Fig. 41 — Fuse panel and converter assembly.

station repeater and bracket details are added to lock the repeater on the slide portion of the bay repeater shelf. The sliding shelf design affords easy access for repeater installation and maintenance.

As with the main station repeater, the manhole A equalizer and associated control equipment were adapted for relay rack mounting. Here, elements of the equalizing repeater are plugged into a deep sliding shelf (Fig. 42) for mounting on the line bays. The lightweight, well braced, welded aluminum shelf carries the ball bearing slides for the movable drawer portion that holds the plug-in units. The welded "hat" section and the bent over end flange which contribute to shelf stiffness are visible in the bottom center of Fig. 42. The spring release latching devices are recessed in the faceplate of the drawer to complement the simple flush front appearance of the shelf assembly. A hinged baffle plate, shown in the open position in Fig. 42, permits access for local cable control, power, and coaxial connections to the plug-in units. For the main station designs, lifting bracket details for easy insertion and removal of the units and locking bracket details for anchoring the units on the sliding shelf are added to the manhole design castings. These units are not coated internally with the high voltage epoxy insulation.

The B equalizer (Fig. 43) is a relay rack mounted aluminum panel-

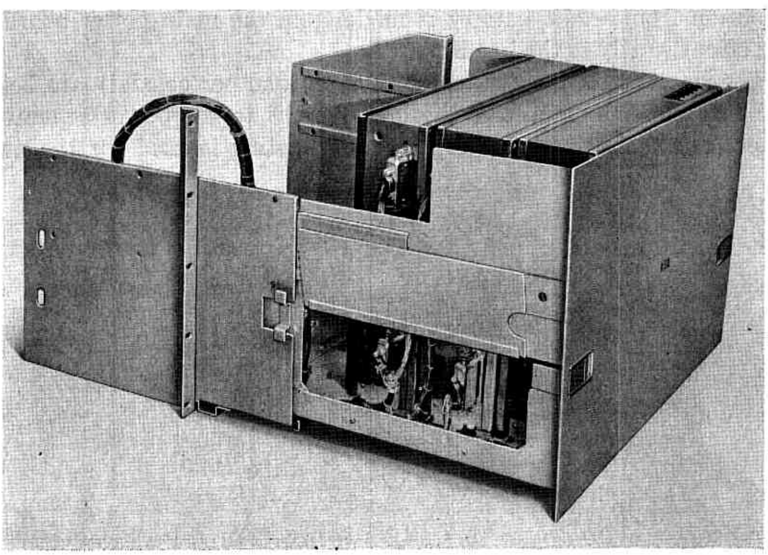

Fig. 42 — A equalizer shelf assembly.

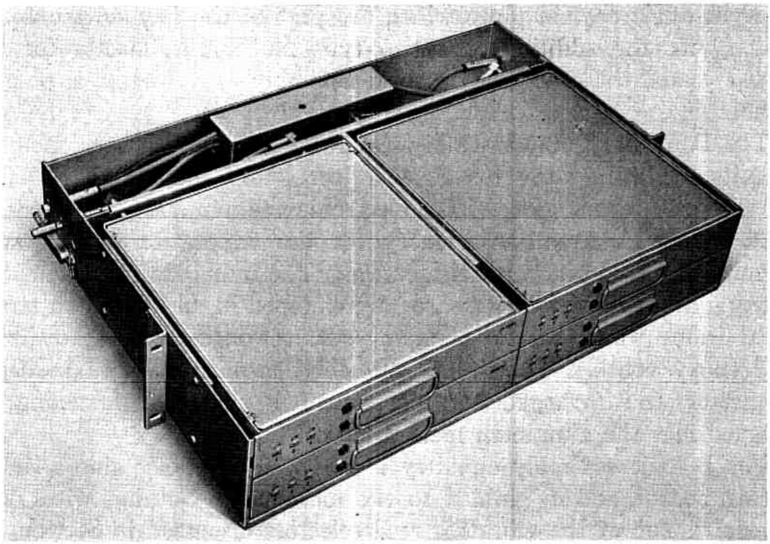

Fig. 43 — B equalizer assembly.

type design. The circuitry is similar to that of the A equalizer and for convenience in performing test and maintenance, the amplifiers, thermistor ovens, oven control networks, and equalizer networks have been divided in an array of four removable, slide-type, plug-in drawers. Small Bell System coaxial plugs, equipped with teflon guide bushings, are rigidly fixed to the drawer and mate with smaller right angle coaxial jacks mounted on floating funnel-type phenolic guide blocks at the base of the B equalizer mounting shelf. Power and control connections are made via a multipin connector assembly which has the necessary float for mating and provides the drawer retaining feature by way of a split spring guide pin arrangement. The amplifiers have their bottom cover welded to the base of the removable drawer.

Equalizer networks, located outside of the amplifier housings, and oven control networks are contained on epoxy glass printed wiring boards mounted on standoffs in the bottoms of the drawers. The top drawers have shields and dust covers which fasten to small floating angle brackets located on the sides of the drawers. Gain adjustments and test points are accessible at the faceplates of the drawers.

The base of the B equalizer mounting shelf houses the shielded combining-hybrid network assembly. Its coaxial leads are terminated in small coaxial jacks mounted at the sides of the shelf for access to right angle coaxial plugs which are part of the bay duct cabling. Thus, with the addition of a lock-type multipin connector for the power leads, the B equalizer shelf assembly is virtually a plug-in panel.

The B memory unit assembly (Fig. 44) contains the ten memory circuits controlling the ten B equalizer bump shapes. This design represents a repackaging of the electronic counter and gate circuitry used in the A equalizer memory for the adaptation of ten such circuits to main station panel mounting. The individual, double sided, epoxy glass printed wiring boards are clamped in an aluminum structural frame which has its clamping surfaces insulated by anodizing and epoxy coating. The frame assembly, terminated in a multipin connector and arranged for mechanical keying and fastening, is plugged into the aluminum memory panel housing.

The B logic assembly circuitry is mounted on epoxy glass printed wiring boards and assembled to the main panel housing with rigid standoffs. Ten of the vibrating reed selectors discussed in Section 3.3 are mounted on a printed wiring board and held by clamps similar to those designed for the A equalizer logic.

4.4.2 *Elements of Line Connecting—Transmitting Bay*

Elements of the L-4 line connecting circuits mounted in the transmitting main station repeater bay consist of the transmitting MMX-2 connecting assembly and the transmitting jack field panel assembly.

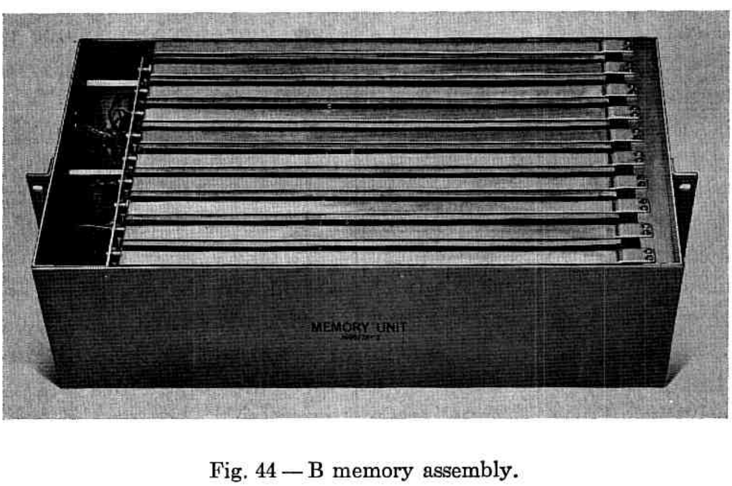

Fig. 44 — B memory assembly.

The former accomplishes the master-group adding function while the latter provides for pilot insertion and the connections from the control center, branching equipment, and receiving equipment.

The transmitting MMX-2 connecting assemblies, one for each coaxial cable, are located below the A equalizer shelves in the transmitting bay (Fig. 39). The MMX-2 connecting unit contains active circuits and is not protected by the line protection switching system. Therefore, this unit has two parallel transmission paths, each of which is monitored by a 512 kHz detector which controls a coaxial switch. The coaxial switch, accessible from the front, is located in the top center of the panel. The switch mounting, a die cast aluminum housing, forms an integral part of the panel assembly and, by virtue of epoxy adhesive bonding, contributes to the structural strength and rigidity of the panel by acting as the main strut in the center span of the assembly.

The active circuits of the two parallel paths are housed in plug-in drawer assemblies inserted from the front on each side of the coaxial switch. There are two sets of drawers consisting of the pilot detector and the pre-emphasis module per set. The drawer fronts have a recessed handle and access to the potentiometer adjustment. The elements within the pilot detector drawer are assembled much in the same way as are the elements within the modules of the MMX-2 Bay.[13] This results in minimum lead lengths between can elements without adverse effect of manufacturing tolerance buildups among the three elements being assembled. Small path selector switches and miniature indicating lamps are mounted in the front face of the main panel.

The center of operational activity on the transmitting bay is the transmitting jack field panel assembly (one per coaxial cable) located at shoulder level in the bay, convenient for people to reach. The hybrid panel and jack field combination contains the line amplifier, the hybrid networks, the line pilot adjust mounting assembly, the combining network printed wiring board assembly, cable equalizer, coaxial connectors, and test access points required for the L-4 line connecting functions.

The shelf provides front face access to the gain adjustment for the line amplifier and level adjustment for the L-4 line pilots. The hybrid networks, on epoxy glass printed wiring boards, are contained in can assemblies for shielding. Connections to the shelf are via small coaxial jacks on both sides of the shelf for signal leads and via a lock-type multipin connector for power leads.

4.4.3 *Receiving Main Station Repeater*

Figure 45 shows the line receiving bay equipment arrangement. The bay contains the repeater and line connecting equipment for one receiving coaxial cable. The type of construction is similar to the transmitting main station repeater bay.

The power separation filter at the top of the bay shares space on its mounting shelf with fault location equipment for main station repeaters as shown in Fig. 46. The oscillator, in the center of the shelf, appears in every other receiving bay and supplies monitoring tones to the main station repeaters of four coaxial cables. The dc-dc converter on the right powers the monitoring oscillator loop adjacent to the main station. Oscillator assembly and converter are the same plug-in units used in the manholes and are adapted for shelf mounting.

The fuse panel is similar to that in the transmitting bay but uses three A buses and one B bus to power the bay equipment. Two A buses power the repeater equipment associated with the one receiving coaxial cable appearing in the bay while the remaining A and B buses feed redundant power for the line connecting equipment which is not protected by the line protection switching system.

As in the transmitting bay, the receiving main station repeater uses the manhole regulating repeater casting with keyed line build-out positions and mounts in the sliding drawer below the fuse panel.

The A and B equalizer equipment is identical to that included in the transmitting bay (Figs. 42, 43, and 44).

Figure 47 shows the band-edge regulator assembly. This type of construction is also used for the monitoring tone blocking shelf and other shelves of the line connecting assembly.

4.4.4 *Elements of Line Connecting—Receiving Bay*

The balance of the panels on the receiving bay comprise the L-4 line connecting, performing mastergroup blocking, branching, and dropping as well as tone blocking, and pilot pickoff, and providing control center receiving connections and access points for in-service testing and maintenance.

The receiving MMX-2 connecting assembly, located below the A equalizer (Fig. 45), is the counterpart of the transmitting MMX-2 connecting assembly discussed earlier and substitutes the de-emphasis module plug-in drawers (featuring the same redundancy and construction) for the pre-emphasis module.

As in the transmitting bay, the center of operational activity on

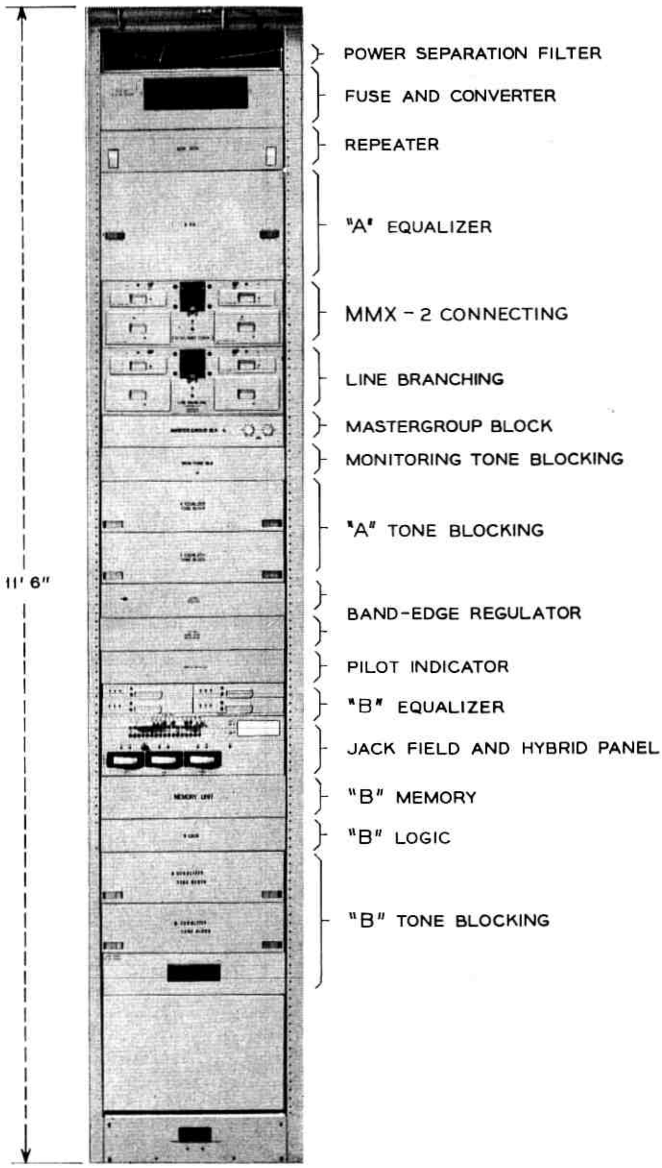

Fig. 45 — Receiving bay equipment arrangement.

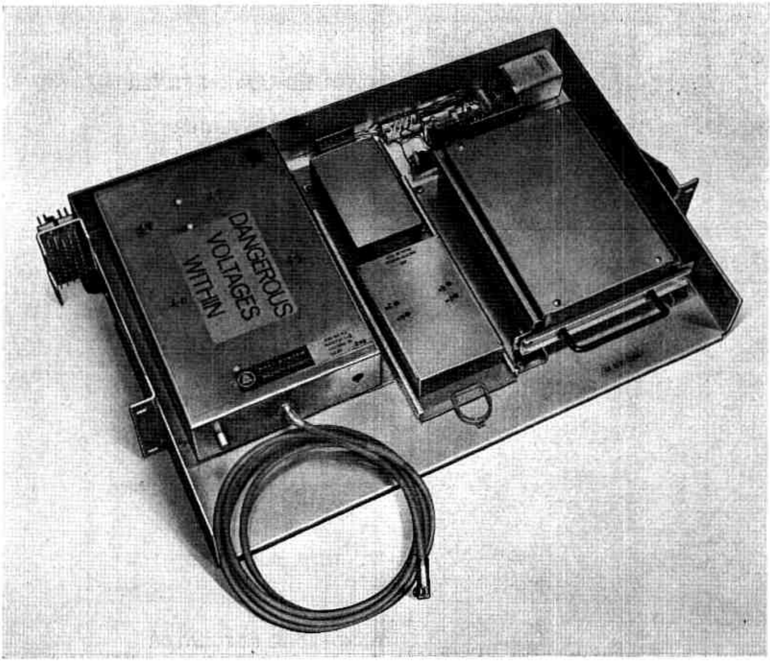

Fig. 46 — Receiving bay—power separation and converter shelf.

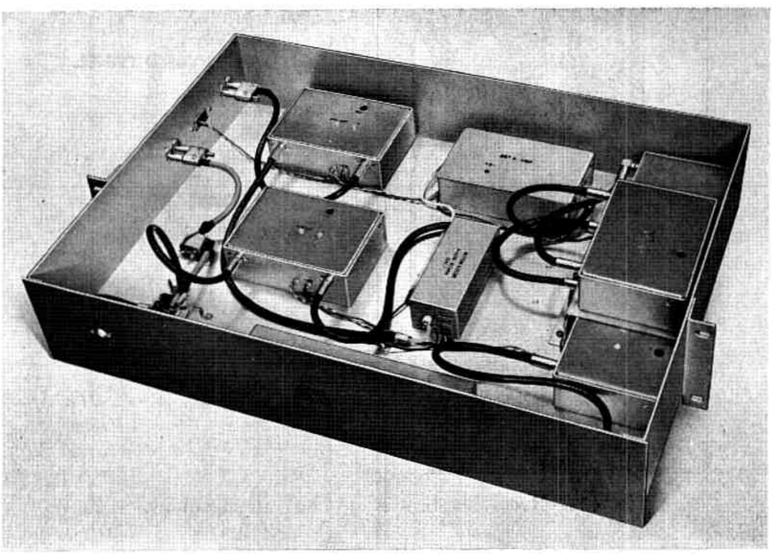

Fig. 47 — Band-edge regulator.

the receiving bay is the receiving jack field and the hybrid and meter panel located at shoulder level in the bay for access. The hybrid panel, in addition to containing the receiving counterparts of the transmitting hybrid panel, includes elements of the line protection switching system switch initiator circuitry. The indicating meters for the continuous L-4 line pilots are mounted in the front of the panel. The line amplifier faces the rear of the panel and gain adjustments are made from the back of the bay.

The intermix of line connecting elements with main station line repeater elements on the bay yields an orderly composite of shelves individually designed to the same ground rules and results in a simple, flush-front, uniform bay appearance which typifies L-4 equipment. Blank shelves and blank panels are available to fill spaces in the bay when circuit or function options eliminate the need for certain equipment. It is intended that packaged bays be shipped from the factory, fully equipped, wired, and tested for ease and efficiency in field installation.

4.4.5 *Control Connecting Bay*

Certain functional elements of the L-4 system are common to all twenty lines in a cable. In the past there has been a tendency to let this type of equipment be handled as miscellaneous in an office. The decision was made early to consolidate this common equipment into an orderly array on a dedicated bay. The L-4 control connecting bay thus fills the roll of the miscellaneous bay, yet offers the many advantages of a shop-assembled, shop-wired, and shop-tested bay. The L-4 bay design philosophy applying to the line bays was followed closely as is illustrated in Fig. 48.

The B test oscillator unit and the test oscillator connector unit are located directly below the fuse panel. The test oscillator unit contains ten plug-in crystal oscillators which are mounted on epoxy-glass printed wiring boards, are enclosed in shielded can assemblies, and are accessible from the rear of the panel. The hybrid combining network is contained on an epoxy-glass printed wiring board mounted on stand-offs in the front part of the shelf cavity. The oscillator connecting unit uses two such hybrid networks similarly mounted in the front of its shelf. These fan out the single test oscillator unit output to two sets of ten coaxial jacks for plug and coaxial cord connection to the B equalizers in the line bays.

The coaxial switches used in the line protection switching system

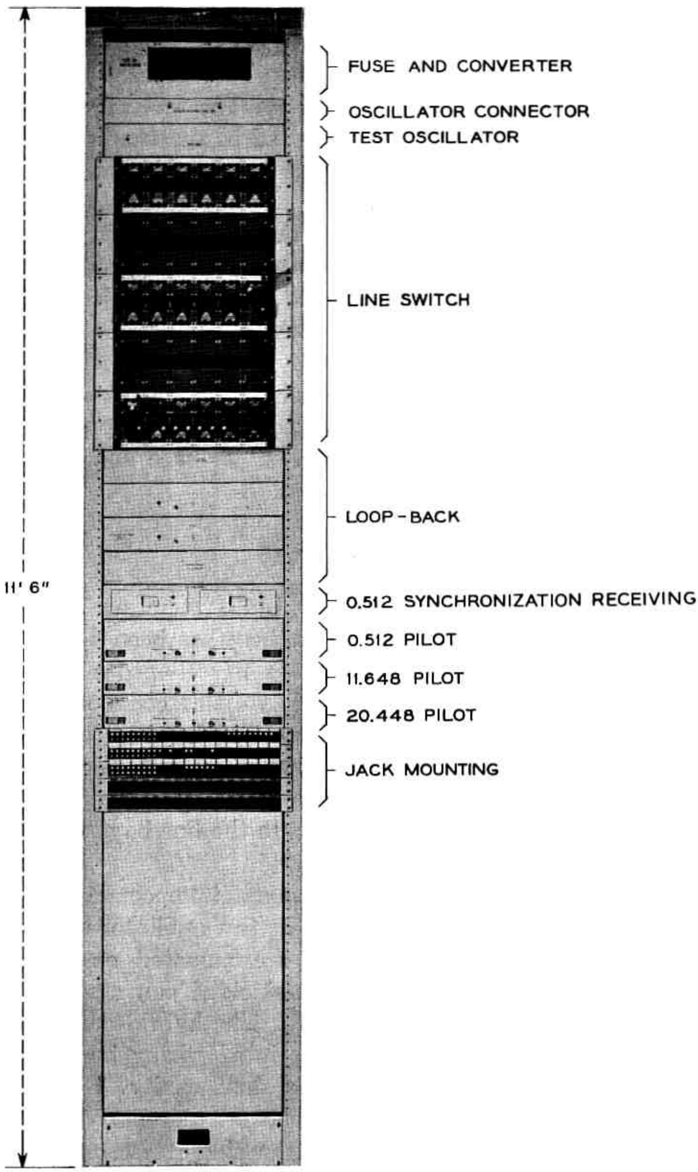

Fig. 48 — Control connecting bay equipment arrangement.

are mounted below the test oscillator and connecting units. The switches plug into die cast aluminum switch mountings which are assembled on panels intended for use in the 19-inch-wide bays used for the L-3 carrier system. In L-3, these switch panels were mounted as miscellaneous.

In L-4, which is initially using a modification of the L-3 line protection switching system, these switch panels are adapted for use on the 23-inch-wide control connecting bay. The jack field associated with the switch array is located at the bottom of the equipped portion of the bay. Again, the jack strips are those used on the L-3 bays and were adapted for use in L-4. Mounted on the jack strips are the indicating lamps, which portray the status of the switches, and the test jacks and switch keys used to check and control the switch modes.

The four panels located below the switch field provide the command looping function of the remote control system and the receiving line connections to the local remote control center. The loop-back portion consists of three shelves which, because of the need for hard wiring among the shelves, are joined on mounting bars to closely associate the individual shelves during the assembly, test, and other operations in the shop. For good control of the wiring within and among the three shelves, local cable designs are used featuring prescribed slack to permit backing off of any one shelf from the rear of the bay for maintenance.

The 512 kHz synchronization receiving assembly is located below the loop-back equipment. It is a four-inch high panel arranged to mount two plug-in modules. The left drawer contains the working circuit; the right drawer is a dead spare conveniently stored in the event that replacement of the working unit becomes necessary. Instantaneous replacement is not a design criterion since the primary frequency generator can run unsynchronized over relatively short intervals, affording adequate time to replace the working module in the event of a trouble. The design principle applying to the drawer modules is similar to that used on the modules of the MMX-2 connecting panels. The input connection from the receiving bay is through a small coaxial jack that mates with a plug connection on the fixed shelf. The connections to the primary frequency generator and battery are via a multipin connector engagement to protect the pins on the module while stored and to hold it in place on the shelf.

Space between the switch jack field and the 512 kHz synchronization receiving assembly is occupied by the L-4 pilot originating equipment. The top unit in this array is the 512 kHz pilot stabilization and distribution assembly. (Fig. 49). This unit receives the 512 kHz pilot from two taps on the distribution bus associated with the primary frequency generator in an office and limits and filters the signals in two independent paths. The outputs of these two paths are fed through differential detector and automatic switching equipment so that failure of the working path will bring about an automatic switch to the standby path.

A distribution network provides ten outputs to feed the ten transmitting coaxials of a route and ten outputs to feed the transmitting MMX-2 connector equipment. The apparatus associated with the dual signal paths is mounted in the sliding drawer. A local cable harness was designed to connect the apparatus to the distribution buses located at the rear of the shelf. Slack in the harness together with appropriate clamping allows the laced cable to twist freely without causing cable flexing and fatigue when moving the sliding drawer. Connections to the distribution bus are hard wired for running into the cable ducts of the bay uprights. Indicating lamps and a test jack position are recessed in the drawer faceplate which also contains miniature rotary switches for manual control and selection of signal paths.

The remaining two panels (Fig. 48) constituting the pilot originating equipment are the 11.648 and 20.448 MHz pilot generator and distribution assemblies. Both units are four inches high and provide dual oscillators with differential detection and automatic switching equipment so that failure of the working oscillator will bring about a switch to the standby. The distribution networks provide ten outputs to feed pilots to the ten transmitting coaxial cables of an L-4 route. The apparatus for these dual pilot sources is housed in the cavities of sliding drawer assemblies as was done with the 512 kHz pilot stabilization and distribution assembly (Fig. 49). An added feature with these pilot generators is the use of plug-in printed wiring board arrangements for the alarm circuits. Small printed wiring boards slide within guide-brackets mounted perpendicular to the main drawer cavity and contain multipin connectors which mate with connectors in the drawer. The L-4 pilot generators resemble the master-group pilot generator designed for the L-4 mastergroup multiplex bay.

Thus, the role of an L-4 miscellaneous bay is effectively filled by the packaged control connecting bay shipped as a unit from the shop. Although the bay presents a wide assortment of equipment which cuts across such various functional entities of the L-4 system as the main repeater, the line protection switching system, the control center

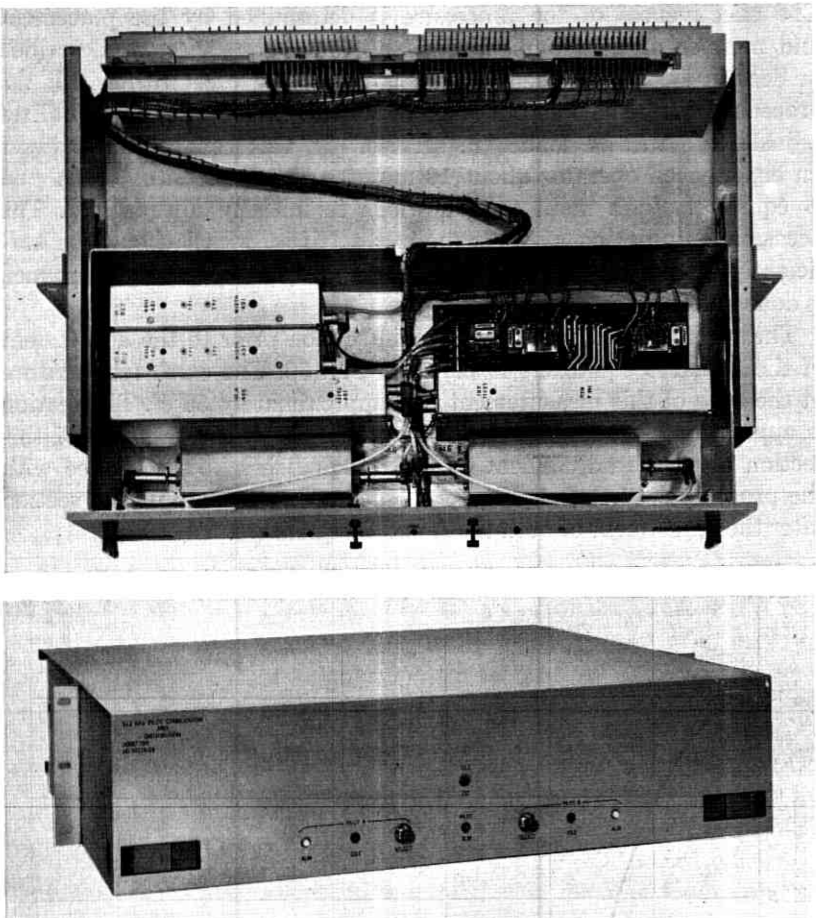

Fig. 49 — 512 kHz pilot stabilization and distribution unit. Top view (top) and front view.

operation, and the pilot generator, the over-all concepts as established for L-4 main station bay physical designs were satisfied.

V. OVER-ALL OPERATION OF THE EQUALIZERS

The procedures followed to determine the proper settings for the A and B equalizers are covered in detail in Ref. 1. It is apparent that the process amounts to adjusting the equalizers so that the response error at the 16 A and B test frequencies is as near 0 dB as possible.

The effectiveness of the procedure is determined by the placement and number of the "bumps" and the test frequencies, as described in Section II. It has been emphasized that the ease with which the procedure can be carried out is an important factor and that equalizer adjustments will be made "in service." In this connection, it takes an experienced operator about 10 minutes to adjust all of the A and B equalizers in a main section which is initially unequalized. The occasional up-dating of the equalizer settings, which should be sufficiently infrequent to place little burden on the operating personnel, is done in a fraction of this time.

The response of a typical L-4 main section prior to the adjustment of the A and B equalizers is shown in Fig. 50. The earth temperature at the time of this measurement was approximately 50°F. The section is approximately 145 miles long. Figure 51 shows the response of this section after the adjustment of the A equalizers in accordance with the procedures of Ref. 1. Figure 52 shows the response of the section after the adjustment of both the A and B equalizers.

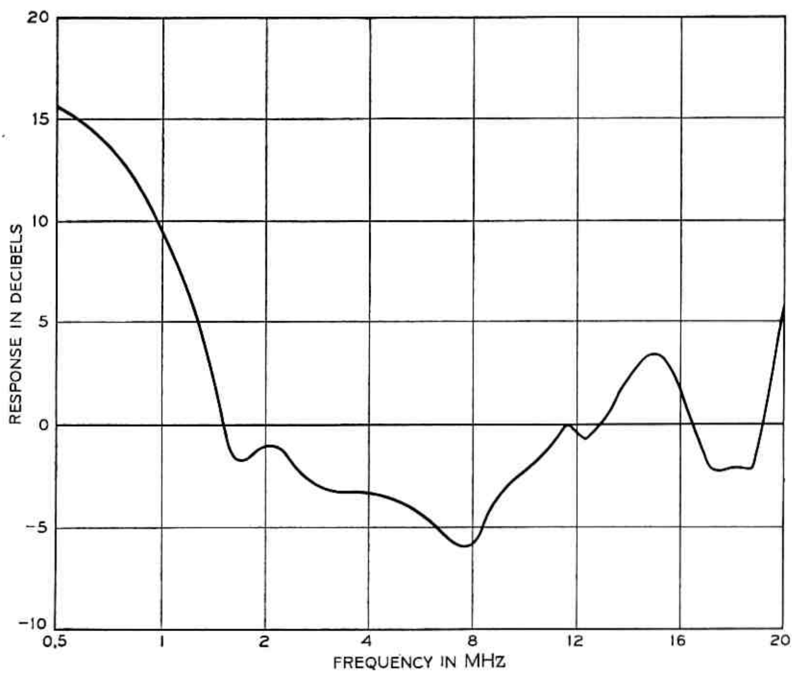

Fig. 50 — Response of L-4 main section before equalizer adjustment.

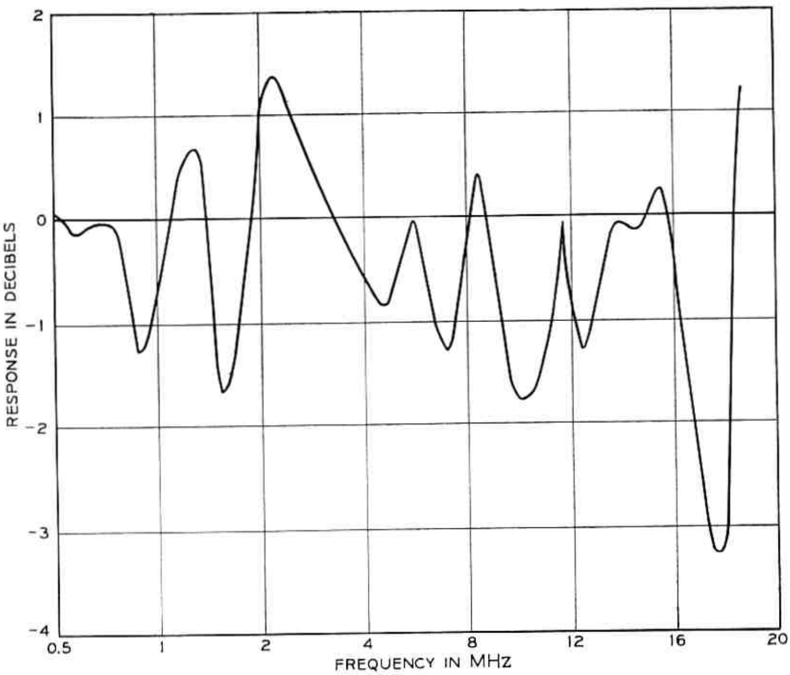

Fig. 51 — Response of L-4 main section after the adjustment of the *A* equalizers.

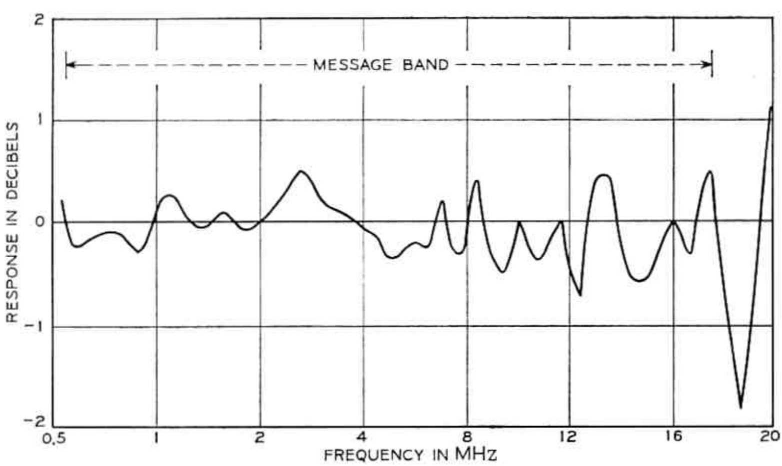

Fig. 52 — Response of L-4 main section after the adjustment of the A and B equalizers.

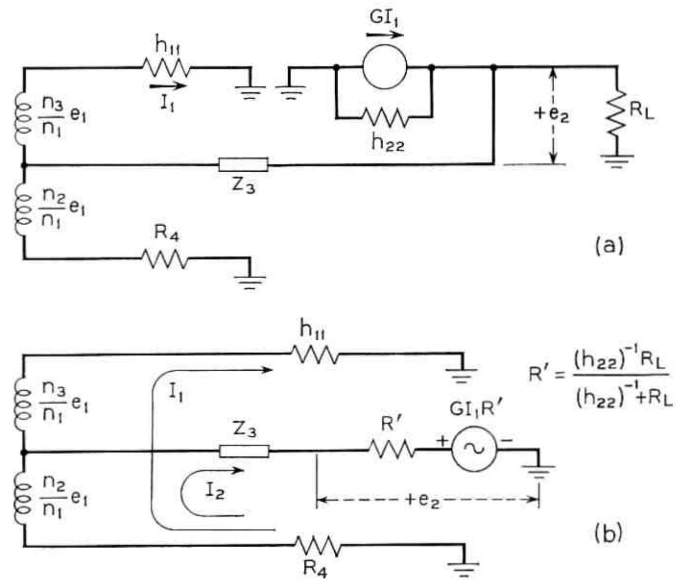

Fig. 53 — Simplified versions of Fig. 7.

APPENDIX

Approximate Gain of the Equalizer Amplifiers

Figure 7 shows the hybrid feedback amplifier configuration used throughout the A and B equalizers. Figures 53(a) and (b) are simplified versions of the circuit of Fig. 7 permitting convenient analysis.

From Fig. 53(a)

$$e_2 = (GI_1 + I_2)\, R'$$

and

$$\begin{bmatrix} \dfrac{n_3 + n_2}{n_1}\, e_1 \\ \\ \dfrac{n_2}{n_1}\, e_1 \end{bmatrix} = \begin{bmatrix} h_{11} + R_4 & R_4 \\ R_4 + GR' & Z_3 + R_4 + R' \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \end{bmatrix}$$

$$I_1 = \frac{\dfrac{n_2 + n_3}{n_1}\, e_1(Z_3 + R_4 + R') - \dfrac{n_2}{n_1}\, e_1 R_4}{(h_{11} + R_4)(Z_3 + R_4 + R') - (R_4 + GR')R_4};$$

$$I_2 = \frac{\frac{n_2}{n_1} e_1(h_{11} + R_4) - \frac{n_2 + n_3}{n_1} e_1(R_4 + GR')}{(h_{11} + R_4)(Z_3 + R_4 + R') - (R_4 + GR')R_4}.$$

Thus

$$\frac{e_2}{e_1} = R' \frac{G\frac{n_2+n_3}{n_1}(Z_3+R_4+R') - G\frac{n_2}{n_1}R_4 + \frac{n_2}{n_1}(h_{11}+R_4) - \frac{n_2+n_3}{n_1}(R_4+GR')}{(h_{11}+R_4)(Z_3+R_4+R') - R_4(R_4+GR')}$$

$$\left.\frac{e_2}{e_1}\right|_{G\to\infty} \to \frac{\frac{n_2 + n_3}{n_1}(Z_3 + R_4) - \frac{n_2}{n_1} R_4}{-R_4}$$

$$= \frac{n_2}{n_1} - \frac{n_2 + n_3}{n_1}\left(1 + \frac{Z_3}{R_4}\right),$$

$$\left|\frac{e_2}{e_1}\right| = \frac{n_3}{n_1}\left(1 + \frac{n_2 + n_3}{n_3}\frac{Z_3}{R_4}\right)$$

$$= \frac{n_3}{n_1}\left[1 + \frac{Z_3}{\frac{n_3}{n_2 + n_3} R_4}\right].$$

Let $n_3 / (n_2 + n_3) R_4 = R_0$; notice that R_0 is equivalent to the parallel combination of the terminations required by windings n_3 and n_2 for hybrid balance if R_4 is the nominal termination for the winding to which it is connected. Then,

$$\frac{e_2}{e_1} = \frac{n_3}{n_1}\left(1 + \frac{Z_3}{R_0}\right).$$

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The Equalizer Remote-Control System

By R. E. ANDERSON and C. R. CRUE

(Manuscript received August 2, 1968)

The equalizer remote-control system performs the role of a master control for the adjustable equalizers in the lines of an L-4 system. A control center located at an attended main station repeater may have jurisdiction over several hundred equalizers, some of which may be as distant as 300 miles. The required equalizer adjustments are determined from the relative powers of signals received from remotely controlled test oscillators at the equalizing locations. When necessary, the equalizers are remotely adjusted in discrete steps by means of commands sent out by the control center. Memory circuits associated with the equalizer networks maintain the network gain settings between adjustments. At each distant main station in a control system, a loop-back unit provides decoding and switching that control equalizers at the location and complete far-end looping paths to the lines to be equalized or checked. The looping paths pass the commands used to address the equalizing repeaters, which are located between the main station repeaters. The control system also comprises part of a fault-location scheme, which includes a monitoring oscillator at each repeater.

I. INTRODUCTION

The equalizer remote-control system has been developed and incorporated into the L-4 carrier system as part of an over-all plan for providing in-service equalization capability. Reference 2 describes the controllable transmission elements, which are step-adjustable equalizers with associated solid-state logic and memory circuits. The equalization control technique is based on measurements of the relative powers from test oscillators when they are successively and remotely connected to the inputs and outputs of the equalizers. Disparities in the relative received powers from different points reveal line misalignment and equalizer settings at the test-oscillator frequencies.

The L-4 control system always consists of a sending and a receiving circuit located at an attended main station repeater and a loop-back circuit at each of one or more remote main stations. The remote locations may be either attended or unattended. All commands originate in the sending circuit; the test signals from the various remote locations are measured in the receiving circuit. The sending and receiving circuits are enclosed in a three-bay console (Fig. 1); the complete assembly is called a control center. The loop-back circuits perform switching and control operations at the remote main stations upon commands from a control center.

The L-4 system also has a built-in fault-location facility for remotely identifying a defective repeater in a line. It consists of a monitoring oscillator at each repeater, and circuits in the control center for remotely turning on groups of the monitoring oscillators and observing the received signals from them.

II. CONTROL SYSTEM PLAN

2.1 *General*

The simplified block diagram in Fig. 2 shows the overall L-4 remote-control system plan and associated nomenclature. In a one-section control arrangement, the adjacent main station repeaters are

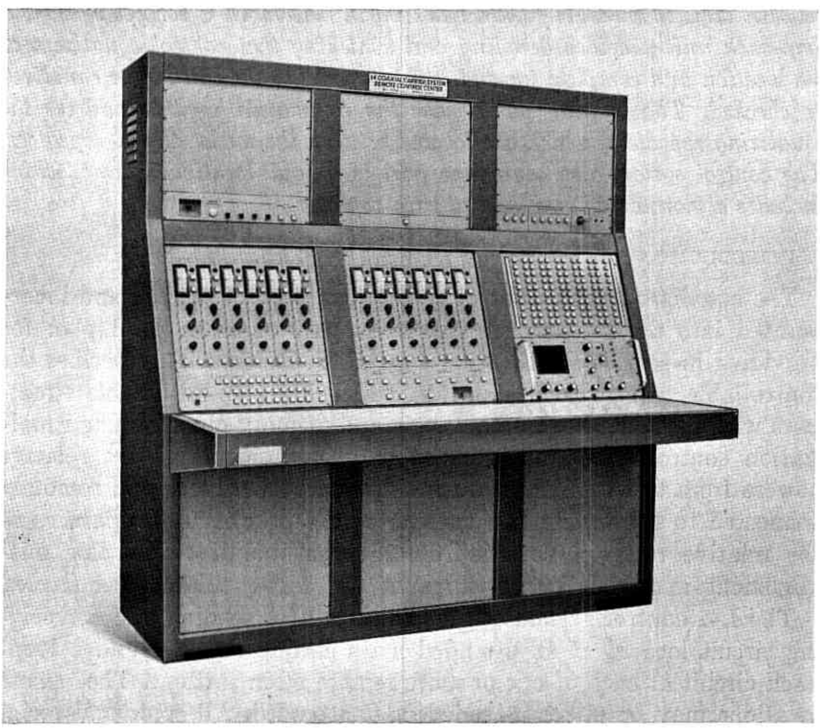

Fig. 1 — Control center.

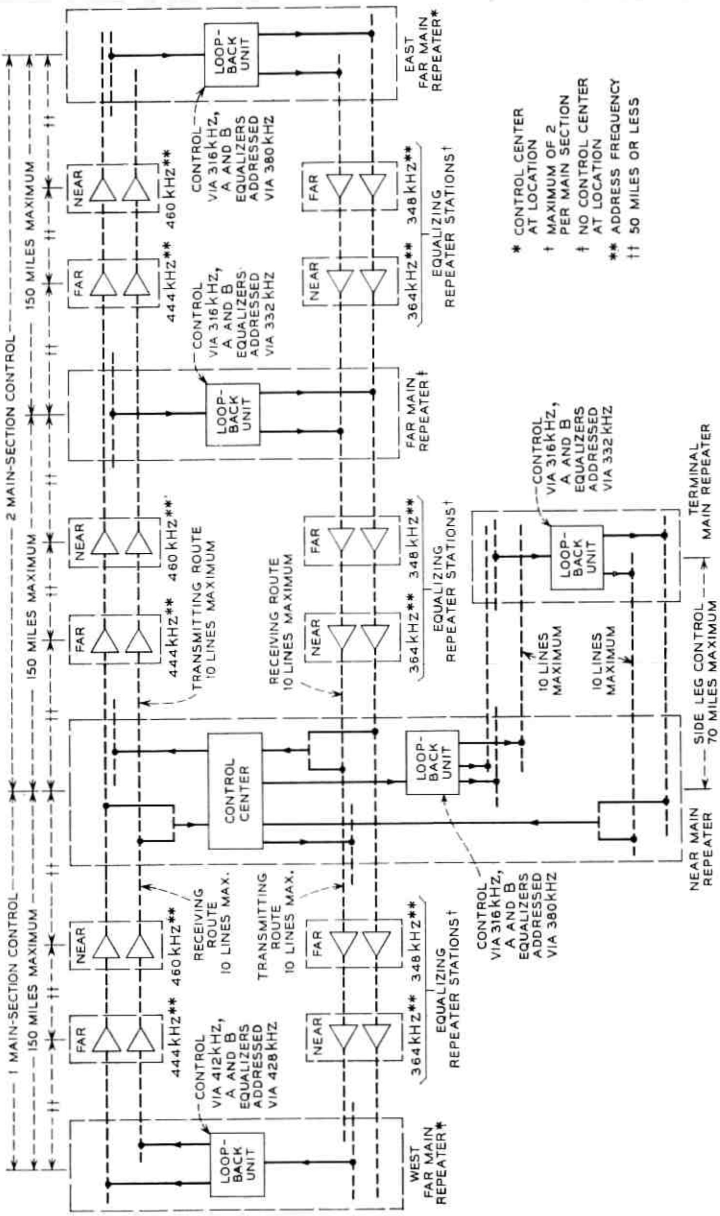

Fig. 2 — Control system—simplified overall plan.

attended and equipped with control centers. In a two-section arrangement, the intermediate main station has no control center, nor blocking, dropping, or branching circuits; it is located between attended locations, each of which has a control center; and it is generally unattended, or at the most, partially attended. One control center may serve up to ten routes; and as suggested by Fig. 2, these may be one-section, two-section, or side-leg arrangements in any combination.

2.2 *One-Main-Section Control*

In the one-section type of operation, the control center at the originating location is used to initiate commands for control operations at the far main repeater, at the near main repeater (originating location), and at up to two equalizing repeater stations between these two locations. The block diagram in Fig. 2 shows a typical layout with two intermediate equalizing repeater stations. The commands for the distant locations are sent out on a regular transmitting L-4 carrier line in the band of 300 to 500 kHz, which is below the lowest message frequencies. The command signals are applied to the line connecting circuit for the regular line on the office side of the automatic protection switch. In this way, the standby line automatically protects the command channel.

2.3 *Commands, the Language of the Control System*

Different combinations of audio-frequency tones are used for the various commands. These are transmitted to a distant location as double-sideband amplitude modulation of a carrier. Envelope detection is then used at the distant location to recover the basic audio command. For control at the originating main station, the audio command tones are transmitted direct to the appropriate A or B equalizer through office wiring. Tuned vibrating-reed selectors are used in the controlled equipment to recognize the command.[3]

As illustrated in Fig. 2, each remote location is addressed by means of a specific carrier frequency assigned to it. A bandpass filter in the controlled equipment passes the appropriate command carrier and the sidebands conveying the audio information.

Test oscillators, provided at each main repeater station and at each equalizing repeater station, are energized and connected to the line during equalizing. Monitoring oscillators, provided at each repeater, are energized for trouble location.

A continuous audio-frequency command tone must be transmitted to keep either type of oscillator turned on. This fail-safe arrangement assures that oscillators will not operate indefinitely with no remote-turn-off capability should trouble occur in the command generating or transmission facilities. Inability to turn off a group of oscillators should not damage equipment but might hamper trouble shooting and interfere with normal work in distant main sections. Monitoring signals are blocked at each main repeater having a control center, but equalizer test signals are blocked only at frogging, dropping, and branching points. Under normal conditions, equalizing work is scheduled so that an operator does not turn on test signals if they would interfere with measurements in another section down the line.

With the present facilities, every command other than those for oscillator turn-on is a burst of approximately 300 milli-seconds. Most consist of a combination of two audio tones. Three tones are the maximum used for any command.

2.4 *Command Carrier Assignment—One-Main-Section Control*

In a typical one-section control system, the west control center generates command carriers in the 300 to 400 kHz band to equalize lines receiving from the east. The east control center (Fig. 2) uses command carriers in the 400 to 500 kHz band to equalize lines receiving from the west. This arrangement provides needed diversity in the command channel frequencies for the equalizing repeaters in the regular line used as a command channel to the distant main repeater. All command carriers are blocked at the adjacent main station to prevent undesired control of distant line sections.

The east control center operates the loop-back unit at the west main station by means of commands transmitted on the 412 kHz carrier. The 428 kHz carrier is used to address the equalizers at that location. After the loop-back unit receives appropriate commands, it routes the 444 and 460 kHz carriers to a desired line for addressing the far- and near-equalizing repeaters. Figure 2 shows all of the command carrier assignments for controlling the eastward transmitting lines and most of the assignments for the westward transmitting lines.

When two equalizing repeaters occur in a main section, the near and far classifications are determined by the direction of transmission and control (Fig. 2). When there is only one equalizing repeater, it is classified as a near-equalizing repeater. More detailed descriptions of the control center and loop-back units are given in later sections.

2.5 *Two-Main-Section Control*

Figure 2 also illustrates a typical two-section system in which control may be extended up to approximately 300 miles. Basically, the loop-back units are similar to those used in a one section system, but a few different features are needed and provided. The command channel frequencies are the same as those used for one-section control, with the exception of those used to address equalizers at the distant attended main station. For example, 380 kHz is used to address the east station instead of 332 kHz as with one section. When control is desired in the near section, the loop-back unit at the unattended location is directed to route commands to the desired line in that section. This loop-back unit also routes commands to the equalizers at the unattended main repeater station. When control is desired in the far section, the loop-back unit at the distant attended main station may be directed to route commands either to equalizers at that location or to desired lines in the far section. As in one-section systems, all command channels are blocked at the attended main stations to prevent undesired control of distant line sections. Also, the upper half of the command band is blocked in the west-to-east lines at the unattended main repeater. The lower half of the command band is blocked in the east-to-west lines. This additional blocking prevents commands that are directed to an equalizing repeater in a far section from simultaneously affecting a similar equalizing repeater in a near section. The pattern of classifying the far and near equalizing repeaters is the same as that for a one-section layout. Figure 2 shows the command-channel frequencies associated with the different locations in the two-section arrangement.

III. SENDING CIRCUITS

3.1 *General Considerations*

The sending circuit, shown in Fig. 3, contains several different individual signal-generating and switching circuits that provide the command originating facilities for the remote control system. Under present arrangements, all commands are initiated manually by operating pushbutton keys associated with some of the switching circuits.

Since high speed is not required in these circuits, miniature relays have been used to advantage in providing economical, noise-immune logic for many of the switching functions. Interlocks have been provided in and among the various individual circuits to prevent generation of false or ambiguous commands and undesired lockups in distant associated loop-back units. In several cases, the interlocks have been designed to force the desired sequences of operating the various controls.

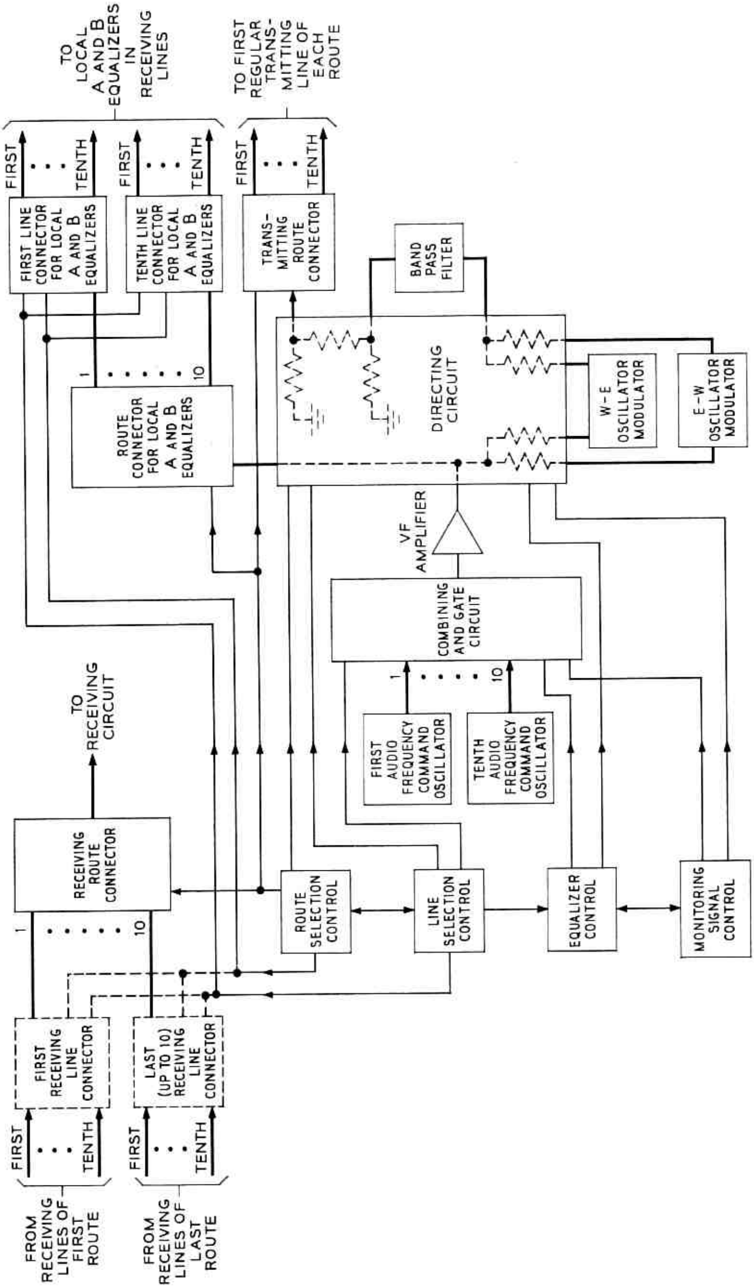

Fig. 3 — Sending circuit block diagram.

3.2 *Route-and Line-Selection Control Circuits*

Route and line selection control circuits are pushbutton-operated switching circuits which use relay and transistor logic to initiate connections to one line section at a time for measurements, equalization, and trouble shooting. The route-selection circuit first selects the desired route; then the line-selection circuit effects connections to a desired line in the chosen route. In one-section operation, the first command addressed to the distant loop-back unit makes the control-system connections to a line at the distant main station. In two-section operation, a choice must first be made between near- or far-section operation. A command initiated by the line selection circuit sets the loop-back unit at the unattended main repeater station for either near- or far-section control. Then the line-selection circuit initiates a line-selection command that is effective in the chosen section.

These circuits also operate the receiving line connecting circuits, which normally associate the receiving facilities of the control center with the line that is receiving command signals. An override control in the line selection circuit provides the means of switching the receiving facilities to a different line of the route. This feature is useful during trouble location and during equalization of out-of-service regular lines in a far main section of two-section control.

Transistor-relay interlocks in the route- and line-selection control circuits prevent the simultaneous selection of two or more lines. The interlocks also preclude releasing or changing a route selection while a line is selected. This prevents leaving operated relays in an idle loop-back unit for an indefinite period.

The line-selection circuit initiates only remote-control commands, namely, those in which the audio commands are conveyed to the controlled point as amplitude modulation of a carrier. A transistor timer determines the command duration. The W-E or E-W oscillator modulator generates the carrier; the audio command oscillators provide the audio tones (Fig. 3). These oscillators, described in sections 3.5 and 3.9, also generate the signals for all other commands that are initiated by the other sending circuits to be discussed. Route selection does not initiate any commands for transmission to a distant location.

3.3 *Equalizer Control Circuit*

After a line has been selected, the equalizer control circuit is operated next, if the equalization is to be checked or adjusted. This pushbutton-operated circuit contains three principal sections of relay and transistor logic. The location-selection part controls the commands that turn on the equalizer test oscillators at various locations; the A or B equalizer selection part initiates the commands and switching for routing subsequent command signals to either the A or B equalizers at the main repeaters; and the equalizer control part initiates the commands for adjusting the equalizer networks.

The location selection or oscillator turn-on command is continuous, whereas a signal burst is used for A or B equalizer selection and equalizer adjustment. Transistor timers set the burst durations. Two different commands are associated with each equalizer network. One increases the gain and the other decreases it. The audio-frequency command tones are transmitted to a remote equalizer as amplitude modulation of a carrier, whose frequency has been assigned to address the location (Fig. 2). The command tones are transmitted directly to the local equalizers through office wiring from the line connectors for local A and B equalizers (Fig. 3).

3.4 *Monitoring Signal Control Circuit*

The monitoring signal control circuit is part of the remote-fault-location facility for the L-4 system. It contains relay-transistor logic and is pushbutton-operated to originate the commands for turning on monitoring oscillators, which are used to identify a faulty repeater from a control center location. Section VII describes the circuit and its relations with other parts of the control system.

3.5 *Audio-Frequency Command Oscillators*

Ten separate plug-in transistor oscillators, originally developed for the *Bellboy®* personal signaling service, provide the audio tones for the various remote-control commands. The oscillators are identical except for a plug-in crystal, which determines the frequency of each. The oscillators operate continuously and are switched in as required by the different sending circuits.

3.6 *Combining and Gate Circuit*

Various combinations of interconnected diode gates provide the means of selecting single tones or combinations of two or three tones from a group of ten for use as remote command signals. When there is no command signal being transmitted, the gates are reverse biased; and the transmission paths for tones are blocked or switched off. During a command, one of the previously described control circuits forward biases a group of gates. This action completes a transmission path from one, two, or three of the audio oscillators.

3.7 *Number of Audio Commands*

The combining and gate circuit provides tone-selection gating for 62 different audio commands. The present control system uses only 52 of these; the other ten are available if more commands are needed for future uses. The required number of audio commands has been minimized by using many of them at several locations. Table I lists the quantity of commands assigned to each class of control function.

3.8 *Directing Circuit*

This circuit, consisting of miniature relays and resistive pads, has three principal functions. First, it routes the audio commands to the oscillator modulators for remote control or to the route connector for local A and B equalizer control at the originating location. Second, it routes keying signals from the originating control circuits either to the W-E or to the E-W oscillator modulator, depending on the transmitting direction of the system route being observed. Third, the circuit provides a simple key-operated relay circuit for disconnecting an attenuator pad from the outgoing path for command signals. Under normal operation, the pad remains in the path. If the command channel gain is abnormally low because of gross misalignment, the pad may be switched out by operating a pushbutton to increase the power of the transmitted command signal about 15 dB.

TABLE I—ASSIGNMENT OF COMMANDS

Quantity	Function
26	Control of distant loop-back operations, such as looping command transmission path to a desired line and routing equalizer adjust commands to A or B equalizers.
2	Control test oscillators associated with *A* or *B* equalizers.
2	Monitoring oscillator control.
1	Terminate two inputs on command-tone combining network during one-tone command.
12	Adjust *A* equalizer networks and first six B equalizer networks.
8	Adjust last four B equalizer networks.
1	Reset all networks in an A or B equalizer to midrange.
10	Unassigned.
62	

3.9 *Oscillator Modulator and Bandpass Filter*

Each oscillator modulator generates one double-sideband, amplitude-modulated signal at a time for use as a remote control command. There are two versions of this unit; one of each is used in each control center. Six different carrier frequencies, spaced at 16 kHz intervals, are available in each oscillator modulator. Under present arrangements, one available carrier from each oscillator is unassigned. As illustrated in Fig. 2, one oscillator modulator is needed for transmitting control signals in one direction; the other is needed for the opposite direction. The carriers generated in this circuit are crystal controlled and modulated by tones from the audio-frequency command oscillators.

The band-pass filter at the output of the modulator passes signals in the frequency band from approximately 300 to 500 kHz and attenuates signals outside of those limits. This filter assures that harmonics from the command carrier oscillators are suppressed sufficiently throughout the message bands on a working L-4 system.

3.10 *Route and Line Connectors*

The route and line connectors are identical 1×10 switching units, each containing 10 miniature wire-spring relays enclosed in individually shielded compartments. The transmitting route connector has one input port, which may be connected to any one of ten output ports. It switches the remote control command signals to an outgoing regular line in an L-4 system route when so directed by the route selection control circuit (Fig. 3).

Transmission is in the opposite direction through the receiving route connector and the receiving line connectors; these units therefore have ten input ports and one output. The receiving route connector switches the output of any one of up to ten receiving line connectors to the receiving circuit in the control center. The receiving route connector is also under control of the route selection control circuit; it operates simultaneously with the transmitting unit. One receiving line connector is associated with each route to switch the output of any one of up to ten L-4 lines to an assigned input on the receiving route connector. The receiving line connector is under control of both the route-selection and the line-selection control circuits.

3.11 *Route and Line Connectors for Local A and B Equalizers*

The route and line connectors for local A and B equalizers are identical 1×10 switching units, each with ten miniature relays assembled on an unshielded printed wiring board. A control center contains one route connector and ten line connectors (Fig. 3). In conjunction with the directing circuit, these units switch audio command tones to the local A and B equalizers in the receiving lines. Each line connector is dedicated to an assigned L-4 system route. These units are operated by the route-selection and the line-selection control circuits, whose interlocks prevent sending command tones to more than one local equalizer at a time.

IV. RECEIVING SECTION

4.1 *Receiving Circuit*

The receiving circuit, shown in Fig. 4, provides: (*i*) adjustable attenuators and indicating meters for measuring A and B equalizer test tones during equalization, (*ii*) a band-pass filter and spectrum analyzer for selecting and displaying received monitoring oscillator tones for fault location, (*iii*) test jacks providing access to the entire received L-4 spectrum and the monitoring tone band.

4.2 *Receiving Amplifiers*

The equalizer tones received from the incoming L-4 lines must be amplified considerably to obtain enough power to deflect indicating meters during measurements and adjustments. Amplifiers of the design used throughout the line connecting and MMX-2 circuits provide the gain needed. In order to minimize the required number of amplifiers, the combined equalizer tones are amplified as much as possible before they are finally separated and amplified individually to drive the indicating meters. In this way, the gain needed in the ten separate tone paths is minimal.

The incoming composite signal must be preamplified to prevent the tone amplitudes from approaching the noise threshold too closely at the input of the first amplifier following the combining network. The two input amplifiers perform this function. The incoming signal contains not only equalizer test tones but also message signals, pilots, and signaling tones. These latter three contribute more energy to the over-all signal than the equalizer test tones and therefore are the controlling factors in establishing the transmission level at the output of the last amplifier in the input group. This level limits the amount of amplification that can be applied to the composite signal as received from a line.

The test tones are separated from the other line signals in the first

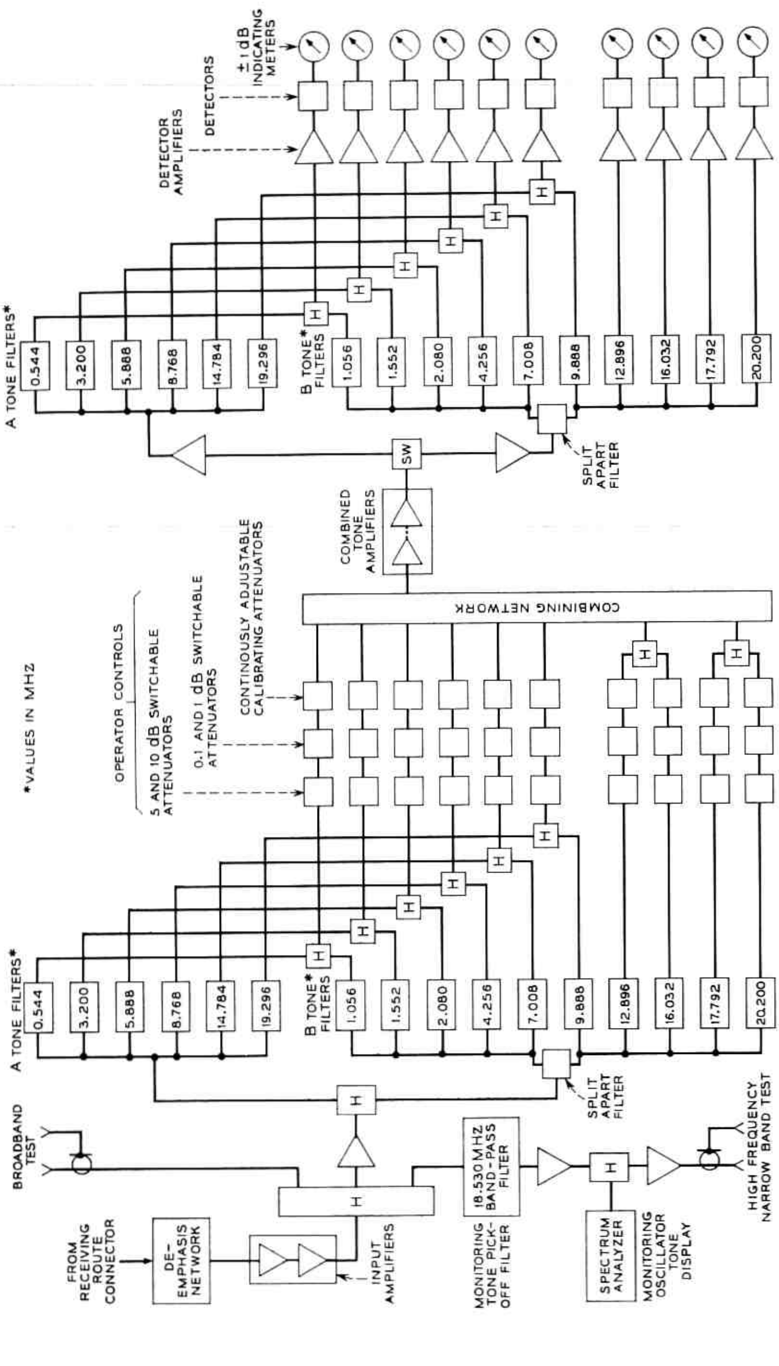

Fig. 4 — Receiving circuit block diagram.

set of crystal band-pass filters. They then pass through adjustable attenuators and are recombined in a resistive network. The new composite signal, consisting of either six A tones or ten B tones, is then passed through a series of tandem-connected amplifiers. This arrangement permits raising the individual tone power much more with combined-tone amplification than would be possible if the other higher energy line signals were also present.

4.3 *Test Signal Power*

The receiving circuit provides a nominal 71 dB of gain to the A and B equalizer test tones. Under this condition, a tone with a power of approximately -63 dBm at the output of the de-emphasis network causes the corresponding detector meter to read zero dB. When the receiving line is not equalized, the test tone power deviates from -63 dBm. Adjustable attenuators with a range of ± 11 dB are provided so that each tone amplitude can be adjusted to null the indicating meters prior to gain measurements. Additional calibrated step attenuators, having a range of ± 5.5 dB in 0.1 dB steps, are used to measure the A and B test signal amplitudes. The indicating meters are calibrated in 0.1 dB increments and have a total range of ± 1 dB. The meters are protected so that they are not damaged when the ± 1 dB range is exceeded.

V. LOOP-BACK UNITS

5.1 *General Description*

A control system always includes a control center and one or more loop-back units, each located at different main stations. All commands transmitted from the control center, except those to equalizers in the same office or to junction loop-back units for side-leg routes, are received by a loop-back unit at a distant main station. The loop-back unit responds to the commands to make a looped connection to transmitting lines for control of A equalizers located at equalizing repeater sites between main stations, or makes direct connections to A and B equalizers within the main station. In addition, loop-back units at attended main stations comprise part of a far-end near-end interlock for control of monitoring oscillators.

5.2 *Main-Station Equalizer Control*

The loop-back units provide common equipment for controlling the equalizers in the main station so that command receiving equipment is not required in each equalizer as at equalizing repeater sites. Connection to the desired equalizer is through miniature relays, which are operated by commands from the distant control center. With this arrangement, a single command receiver can serve up to ten A and ten B equalizers at an attended main station and up to twice as many at an unattended main station. Half of the unattended equalizers are part of the receiving line equipment in the far section of the control system, and half are part of the transmitting line equipment in the near section.

All main station A and B equalizers are controlled by audio-frequency command tones from either a control center or a loop-back unit. Thus, all main station equalizers are alike for both local and remote control. This facilitates maintenance and administration of the equalizers.

5.3 *Loop-Back Unit Types*

There are several types of loop-back units. The particular one installed at a main station depends upon whether the main station is unattended, attended with one- or two-section control, or is part of a side-leg route. Figs. 5 and 6 are block diagrams of two loop-back units. Fig. 5 is a loop-back unit for an unattended main station, and Fig. 6 is for an attended main station where control extends over two main sections. These are the units required in the two-main-section control system illustrated in Fig. 2.

5.4 *Input Amplifiers and Hybrid Network*

The input amplifiers receive the signals at a very low level from the output of the automatic switch for the first operating line and raise the signal amplitude as required for proper operation of the individual loop-back circuits. The amplifiers are broadband and are the same type as used in the control center receiving circuit and elsewhere in the L-4 system. The hybrid network splits the incoming signal path to connect to the individual loop-back circuits.

5.5 *Equalizer and Control Command Receivers*

The equalizer and control command receivers demodulate the equalizer and control command carrier frequencies. The circuits are the same as those in the command receiver located in the line equalizing repeater.[2] Each command receiver contains a narrow band-pass filter, which removes all message and tones from the composite received signal and passes only the appropriate command carrier and

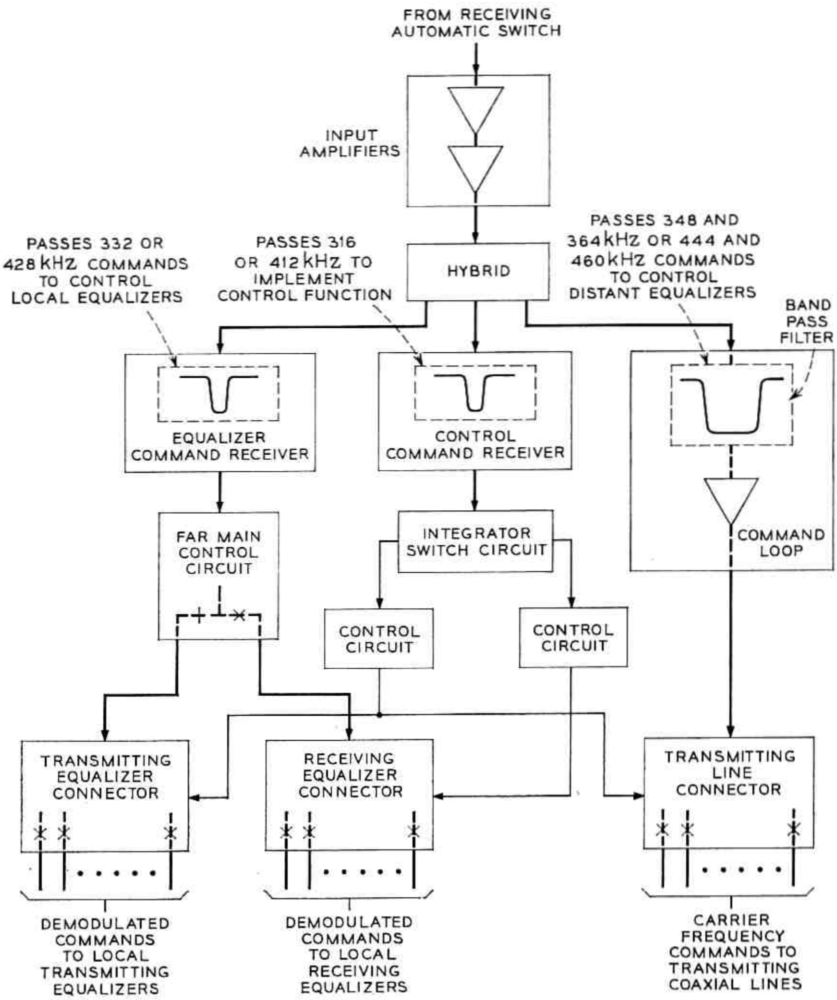

Fig. 5 — Loop-back circuit for unattended main station.

sidebands. The carrier frequency is amplified and detected so that the audio-frequency command tones are obtained at the output.

The command receivers have a gain of approximately 60 dB so that the audio-frequency commands at the output have sufficient amplitude to drive tuned vibrating reed selectors used to separate the command tones.

5.6 *Integrator Switch Circuit*

The integrator switch circuit receives command signals from the control command receiver as bursts of audio-frequency tones, separates the individual tones by means of vibrating reed selectors, and converts them into dc pulses with transistor gates. Most control commands consist of a combination of two audio tones so that two con-

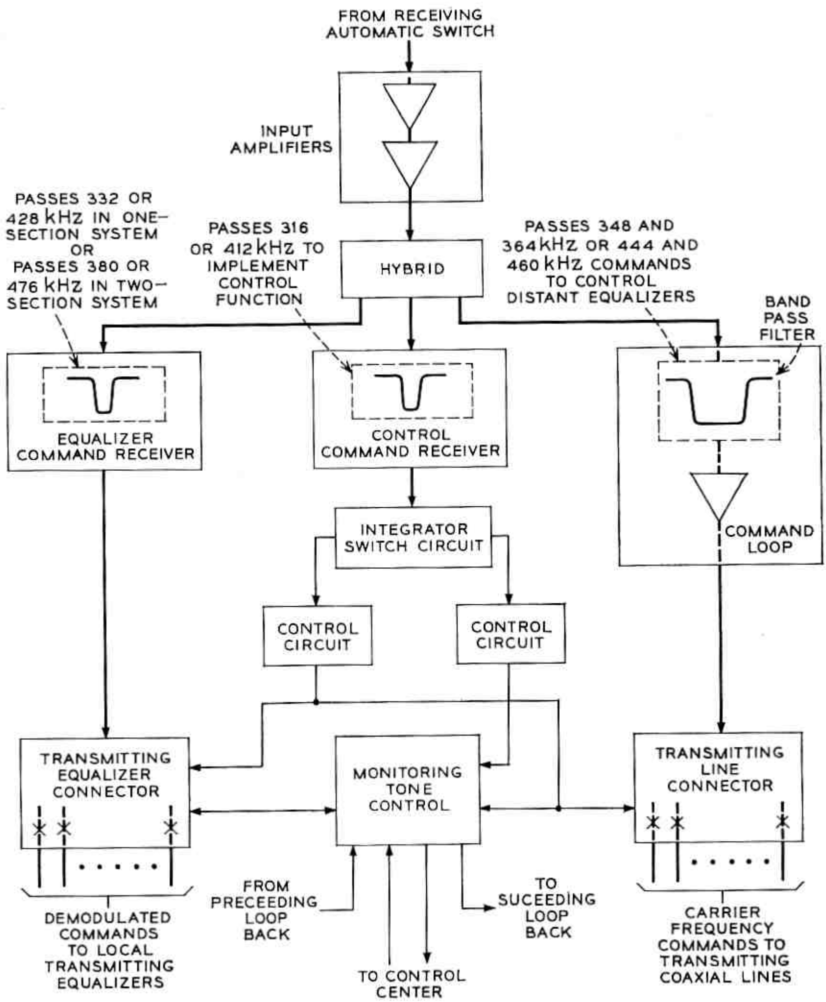

Fig. 6 — Loop-back circuit for attended main station (two main-section control).

current dc pulses are obtained. The duration of the pulses is approximately the same as the duration of tone burst.

5.7 *Control Circuit*

Control circuits respond to dc pulse combinations from the integrator switch circuit to implement line and A or B equalizer connections. In addition, at attended main stations where control centers are present, the control circuit comprises part of the far-end near-end monitoring oscillator interlock between the local control center and the loop-back circuit.

Loop-back units, which are part of control systems extending over two main sections, have two control circuits. One is associated with near-section control functions and the other with the far section. For instance, in the loop-back unit for the unattended main station (Fig. 5), the control circuit at the left operates the line and transmitting AB equalizer connectors, both of which are part of the near section in the control system. The control circuit at the right responds to far-section commands for control of the receiving A and B equalizers.

Conversely, in the loop-back circuit for the attended main station (Fig. 6), the control circuit at the left responds to far-section commands, while the one at the right responds only to near-section line-select commands for inhibiting monitoring oscillator control.

All of the control circuits are identical, and they respond to near- or far-section commands, depending on how they are connected to the integrator switch circuit. The circuit has logic in the form of transistor gates and provides memory by a relay lockup.

5.8 *AB Equalizer Connector*

The AB equalizer connector makes the connection between local A and B equalizers and the equalizer command receiver. There are transmitting and receiving AB equalizer connectors in loopback units for unattended and side-leg terminal main stations. The transmitting and receiving circuits are identical. The transmitting unit makes connections to local A and B equalizers transmitting away from the main station; the receiving unit connects to the equalizers in the receiving lines.

The circuit is comprised of miniature relays which make the connections between the equalizer command receiver, the equalizers, and an AB select circuit. The AB select circuit is a relay transfer switch controlled by transistor gates. The gates are operated in response to A or B equalizer select commands.

Operation of the circuit is such that, normally, connection is made to the A equalizer in the selected line. If it is desired to send commands to a B equalizer, operation of the B select key at the distant control center generates a control command that results in operation of the AB transfer relay. The relay locks up, and the B equalizer remains connected until the A equalizer in that line is selected, or until the line is released or a new line selected.

5.9 *Command Loop*

Commands being transmitted to equalizing repeater sites between main stations are connected to the transmitting line through the command loop, which includes the command band-pass filter and loop amplifier. The command band-pass filter blocks all of the undesired command carrier frequencies and message. The loop amplifier provides gain so that commands are connected to the transmitting lines at the desired power. The transmitting line connector is comprised of wire spring relays which connect the command loop to the desired coaxial line. The transmitting line connector also has one input port which can be connected to any of ten output ports upon command from the distant control center.

5.10 *Far-Main Control Circuit*

The far-main control circuit is required in those loop-back units that are connected to A and B equalizers in both the transmitting and receiving lines in the main station. The far-main control circuit connects the output of the equalizer command receiver either to the receiving or to the transmitting AB equalizer connectors. It consists of a transfer relay controlled by transistor gates and is similar to the AB select circuit in the AB equalizer connectors. Near-section, far-section control commands from the control center operate the gates to control the relay. Normally, the relay is released and connection is made to the transmitting AB equalizer connector. When control is exercised over far-section equalizers, the relay is locked up. It is released when the near-section is selected or lines are released at the control center.

5.11 *Monitoring Tone Control Circuit*

The monitoring tone control circuit is part of the far-end near-end interlock for control of monitoring oscillators. This circuit is required in loop-back units at attended main stations where there are control centers. It consists primarily of an inhibit circuit interconnected with line-select relays in the AB equalizer control circuit.

The inhibit circuit includes transistor logic operated by signals from the integrator switch circuit and a timer. The gates and timer control an inhibit relay, which is part of the monitoring oscillator control interlock with the control center. The overall monitoring oscillator control arrangement is described in following sections on fault location and the far-end near-end interlock.

VI. EQUALIZER ADJUSTMENT

6.1 *General Discussion*

Adjustable networks in the A and B equalizers function in specified frequency bands to minimize the system gain deviations remaining after operation of the regulating repeaters.[2] The A equalizers provide the initial, relatively coarse, gain bump corrections; the B equalizers follow with the final, comparatively finer, bump corrections.

The set of test frequency oscillators associated with the equalizer networks in any given equalizer produces the same output power (within 0.15 dB) and frequencies as those in any other equalizer. This permits gain deviation measurements between any two points in terms of the difference in measured power from the two points.

Interruption of dc power to an equalizer causes the settings of the memory elements to be lost, and random settings appear when power is restored. To expedite re-equalization, random memory settings can be cleared and all of the gain shapes reset to their midrange positions by means of a single command from the control center.

6.2 *Theory of Equalizer Adjustment*

Figure 7(a) shows the A equalizers in one main section, which may be up to 150 miles long. The location designations for the equalizers are:

FM—far main repeater,
FE—far equalizing repeater,
NE—near equalizing repeater,
NM—near main repeater.

These designations are the same as those used on the equalizer selection and control panel of the control center. The panel is part of the equalizer control circuit discussed in Section 3.3.

The test-frequency oscillators, shown symbolically in Fig. 7, are outside of the equalizers to indicate that their outputs can be switched

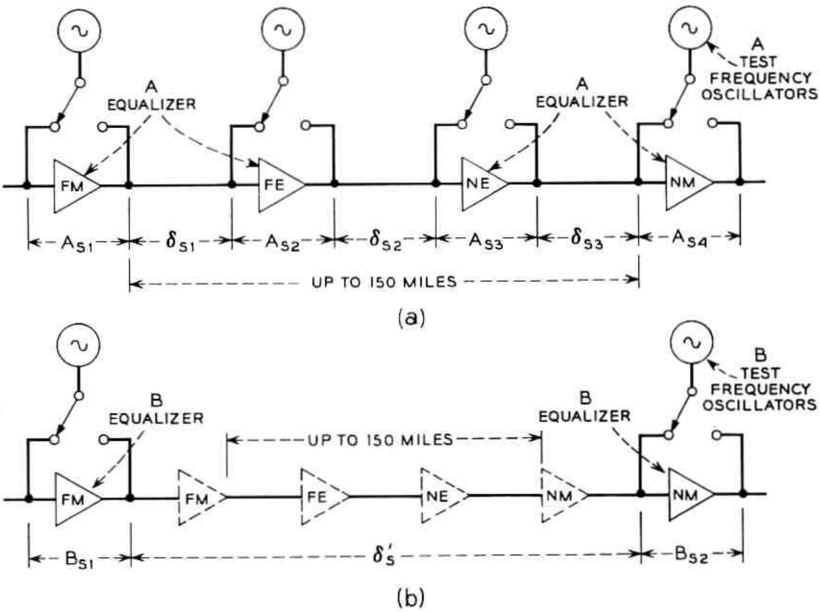

Fig. 7 — (a) A equalization; (b) B equalization.

either to the input or output. This is accomplished by operating appropriate keys on the control panel.

Figure 7(b) shows the B equalizers located at the two main stations. The test-frequency oscillators for these equalizers are also controlled by keys on the equalizer selection and control panel. The following definitions, as indicated in Fig. 7(a), will be used to describe the A-equalization procedure.

$A_{s\ell}$ = A equalizer gain setting at the test frequency associated with the Sth adjustable equalizer network at the ℓth equalizer.
δ_{sk} = gain of the kth line section at the Sth test frequency.

Since there are six A equalizer networks, S can be any integer from 1 to 6.

The values of δ_{sk}, as they deviate from zero, represent line section misalignment, which is corrected for by making compensating changes in the equalizer gain setting $A_{s\ell}$. The gain of the section at a particular test frequency may be expressed as

$$M_{SA} = A_{S1} + \delta_{S1} + A_{S2} + \delta_{S2} + A_{S3} + \delta_{S3} + A_{S4} . \tag{1}$$

As discussed in Ref. 2, pre- and postequalization are used for each line section to minimize modulation and noise penalties on the system performance; thus each equalizer corrects for one half of the deviation in the adjacent line sections.

Therefore, (1) can be conveniently rearranged as

$$M_{SA} = \left[A_{S1} + \frac{\delta_{S1}}{2}\right] + \left[\frac{\delta_{S1}}{2} + A_{S2} + \frac{\delta_{S2}}{2}\right] + \left[\frac{\delta_{S2}}{2} + A_{S3} + \frac{\delta_{S3}}{2}\right] + \left[A_{S4} + \frac{\delta_{S3}}{2}\right]. \qquad (2)$$

When the equalizers are properly set, each bracketed term in (2) is zero; and the section misalignment, M_{SA}, is zero. If the main section equalized, the required changes in equalizer gains are

$$\begin{aligned} \Delta A_{S1} &= -\left[A_{S1} + \frac{\delta_{S1}}{2}\right] \\ \Delta A_{S2} &= -\left[A_{S2} + \frac{\delta_{S1} + \delta_{S2}}{2}\right] \\ \Delta A_{S3} &= -\left[A_{S3} + \frac{\delta_{S2} + \delta_{S3}}{2}\right] \\ \Delta A_{S4} &= -\left[A_{S4} + \frac{\delta_{S3}}{2}\right]. \end{aligned} \qquad (3)$$

The gain changes are implemented by transmitting equalizer adjust commands to the several equalizers. The gain corrections given by (3) can be translated into the number of commands required to effect the correction. Let

k_A = change in A equalizer gain per step

and

$n_{s\ell}$ = number of times the (+) or (−) keys must be depressed for correction of the sth shape at the ℓth location.

Then

$$\begin{aligned} n_{s1} &= \frac{1}{k_A}\,\Delta A_{s1}, \qquad n_{s2} = \frac{1}{k_A}\,\Delta A_{s2} \\ n_{s3} &= \frac{1}{k_A}\,\Delta A_{s3}, \qquad n_{s4} = \frac{1}{k_A}\,\Delta A_{s4}. \end{aligned} \qquad (4)$$

Normally, the system gain deviation and the values of $n_{s\ell}$ are small. The section misalignment remaining, after A equalization, is removed by B equalization. In Fig. 7(b), the section misalignment to be corrected is defined as δ'_s where

δ'_s = gain deviation at the sth B test frequency.

The B equalizers have ten adjustable networks, so s can be any integer from 1 to 10. The equalizer gains are defined as $B_{s\ell}$ where

$B_{s\ell}$ = B equalizer gain setting at the test frequency associated with the sth adjustable equalizer network at the ℓth equalizer.

The section misalignment at B test frequencies is then given by

$$M_{sB} = B_{s1} + \delta'_s + B_{s2} . \tag{5}$$

As is the case for an A equalization, B_{s1}, B_{s2} and δ_s are obtained by measurements at the B test frequencies; pre- and postequalization are used so that each B equalizer corrects for one half of the line-section misalignment. Hence, the required corrections in the B equalizer gains are

$$\Delta B_{s1} = -\left[B_{s1} + \frac{\delta'_s}{2}\right], \qquad \Delta B_{s2} = -\left[B_{s2} + \frac{\delta'_s}{2}\right]. \tag{6}$$

The required gain corrections can be translated into the number of commands required as

$$n'_{s1} = \frac{1}{k_B}\,\Delta B_{s1}\,, \qquad n'_{s2} = \frac{1}{k_B}\,\Delta B_{s2}\,, \tag{7}$$

where k_B is the change in B equalizer gain per step, and n'_{s_1} and n'_{s2} represent the number of times the (+) or (−) keys must be depressed to adjust the sth network in the FM and NM B equalizers, respectively.

The values of $n_{s\ell}$ and $n'_{s\ell}$ to correct A and B equalizer gain settings are accurate only to the extent that k_A and k_B are constants. This condition is not precisely met in the equalizers, although for a given gain shape, the k_A and k_B can be assumed constant. k is approximately 0.3 dB for A equalizer gain shapes and 0.2 dB for B equalizer gain shapes. Any resulting errors in equalizer gain can be corrected by touching up the gains of the A and B equalizers at the main stations.

Another factor that must be considered, when applying equalizer corrections, is the interaction between equalizer networks. Since some of the gain shapes overlap, it is required that the adjustments be made in a sequence that minimizes the interaction effects. In the sequence followed, a group of noninteracting shapes is adjusted first. After these shape changes have stabilized, a second group is adjusted. The shapes in the second group are affected by the first but do not react on the first group or each other. This procedure eliminates the need for additional measurements and computations or automatic computer networks to correct for the interactions.

The (+) and (−) keys, located to the right of each meter, are used to send commands to adjust the gain shape associated with that section of panel. Each time the (+) key is depressed, the equalizer gain is increased one step.

VII. FAULT-LOCATION SCHEME

7.1 *General Description*

A testing arrangement has been designed and built into the L-4 system to provide a means of remotely identifying a faulty repeater. The arrangement is also useful for quickly verifying, under apparently normal conditions, that the gains of individual repeaters are in fact approximately normal. Utility of this testing facility is limited during certain kinds of total failures and during the type of transmission impairment, such as excessive intermodulation, in which there may not be significant gain abnormality.

As part of the fault-locating equipment, each repeater location contains a monitoring oscillator. Upon command from a control center, one group of these oscillators may be turned on at a time in the lines and routes served by the control center. Power is supplied to the oscillators by the main and equalizing repeater stations.[2] Each group contains about half of the oscillators between adjacent power supply points. For example, a group powered from a main repeater station covers about half of the repeaters leading to the adjacent equalizing repeater. Similarly, the groups on either side of an equalizing repeater cover about half of the repeaters in the respective, adjacent line sections.

Each oscillator in a group emits 1 of 16 frequencies spaced at 4 kHz intervals between 18.500 and 18.560 MHz. In each group, the frequencies are assigned consecutively to the repeaters to facilitate identification in case of trouble.

The command for turning on a group of monitoring oscillators is initiated in the monitoring signal control circuit, part of the sending section of the control center (Fig. 3). Any one of up to six groups of oscillators may be selected in a line section by depressing the appropriate pushbutton key on the rightmost sloping front panel of the control center (Fig. 1). Interlocks among this and other circuits in the control center prevent generating a monitoring signal turn-on command until after a route and line have been selected and the equalizer control circuit has been released. The turn-on command is directed to an A equalizer at the location that energizes the group. Logic and switching associated with the A equalizer then complete the power connection.

7.2 *Local Interlock*

The transistor interlock in the monitoring signal control circuit prevents turning on two groups of monitoring oscillators at the same time. If two or more of the six pushbuttons are depressed simultaneously, all of the relays for initiating turn-on commands will remain released until only one is depressed. Control of monitoring oscillators may be released by depressing the release pushbutton, by changing the line selection, or by releasing the line selection.

7.3 *Received Display*

A spectrum analyzer in the control center displays the received monitoring signals as a group of evenly spaced pips. In a normal system, the envelope outlined by the tops of the pips is smooth for oscillators located between regulating repeaters, as shown in Fig. 8. However, there may be a tilt to the display if the cable is not at its annual mean temperature. Under this condition, the regulating repeaters insert gain corrections so that small discontinuities appear at points in the display corresponding to where the regulating repeaters are located. A trouble is indicated by a large amplitude difference in adjacent pips or by missing pips. A photographic record is kept of the normal display for each group of oscillators to facilitate picking out a trouble. If an oscillator should fail, it can be differentiated from a repeater failure since this results in the disappearance of a single pip in an otherwise normal display.

7.4 *Far-End Near-End Interlock for Monitoring Signal Control*

When a group of monitoring oscillators is turned on, monitoring signals are applied to four lines, two in each direction. Interlocks pre-

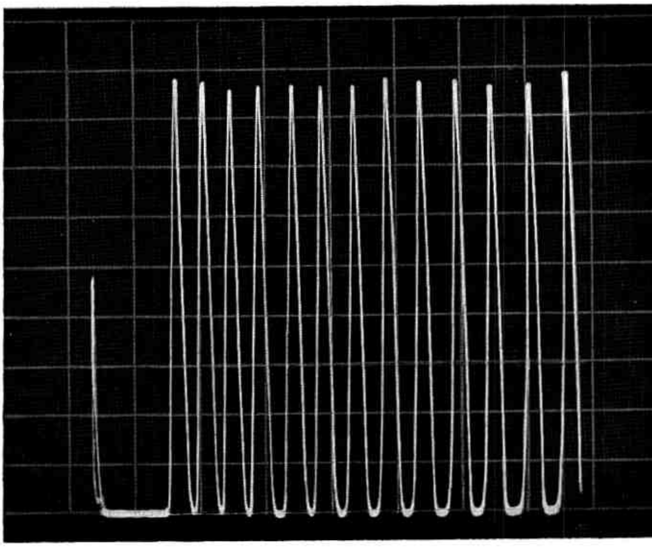

Fig. 8 — Typical spectrum analyzer display of received monitoring tones.

vent the control centers at both ends of a control section from turning on oscillators in the same lines at the same time. This precludes the confusion that would occur if two groups of monitoring oscillators are turned on simultaneously in a line section. The confusion would result because the spectrum analyzer display would include multiple signals at the same nominal frequencies.

Before a command can be transmitted to turn on a group of monitoring oscillators, the conditions imposed by the far-end near-end interlock, as well as those imposed by the local interlocks, must be met. The far-end near-end interlock is arranged between the sending circuit of a control center and the loop-back units at that location. Each of the loop-back units is associated with a different route under jurisdiction of its own distant control center.

Each loop-back unit receives route- and line-select information from the local control center, and monitoring-tone and line-select information from the distant control center. This information is then used in the loop-back units to complete or interrupt a circuit path between the line-select and monitoring-tone control circuits in the control center. When the circuit path is opened, the monitoring tone controls associated with the appropriate route and particular pair of lines become inoperative. Under this condition, the operator cannot turn on any group of oscillators on that route and those lines; and lighted lamps indicate that the distant control center operator has turned on monitoring oscillators.

Although the local operator is inhibited on one pair of lines, he can still observe the tone group that has been actuated on these lines by the distant location. He can also turn on and observe monitoring tones in any of the other pairs of lines.

Once inhibited, the local control center can turn on monitoring oscillators only if the attendent at the distant control center does any one of the following: (*i*) turns off monitoring signals, (*ii*) changes line selection, or (*iii*) releases line selection.

VIII. POWER AND FUSING

An operating control center draws about 10 amperes of current from the unregulated 24-volt office battery. Three dc-to-dc converters in the control center console are energized from the office battery to provide regulated sources of 25-volt power for the various control circuits and amplifiers. Some of the front panel indicating lamps operate direct from the office battery. The 24-volt and 25-volt power is distributed to the control circuits through standard alarm-type fuses, which are located in the console.

The spectrum analyzer operates from the office 115-volt, 60-cycle, ac power. The loop-back units are energized from other dc-to-dc converters, also operated from 24-volt office battery. These are located in the control-connecting bay, which is associated with the L-4 line equipment in the main station. A typical operating loop-back unit causes a current drain of about 1.3 amperes on the office battery. Fuses in the control-connecting bay distribute the regulated 25-volt power to the various amplifier and control circuits that form the loop-back unit.

The fuses for both the control center and the loop-backs provide overcurrent protection and the means of removing voltage to the circuits for maintenance. A blown fuse in either of the groups operates an alarm relay, which in turn activates the office alarm system.

IX. EQUIPMENT ARRANGEMENTS

The control center is assembled in an enclosed three-bay console having both vertical and sloping front panels and a full-length writing shelf (Fig. 1). The sloping front panels contain most of the facilities that are touched or observed by an operator to initiate commands, to make measurements, and to locate remote faults. These front-panel facilities consist of pushbutton keys, indicating lamps, knob-controlled adjustable attenuators, meters, and a spectrum analyzer. Lighted lamps identify the selections of the route, line, equalizing location, and group of turned-on monitoring oscillators. The two meters at the extreme right and the attenuators directly below are spare equipment.

The various control circuits located inside the console are assembled on printed wiring boards. In general, each board is dedicated only to one basic function, such as route or line selection. Local cables provide the connections among the various control boards, and between the boards and the operating keys and lamps on the front panels.

The printed wiring boards are mounted on shelves that are accessible by opening the doors on the back of the console. Those shelves containing active devices and relays are equipped with slides to facilitate quick pull-out for circuit maintenance. The other shelves that contain only passive elements are screwed to the flanges of the uprights. Loops in the local cables permit limited pull-out of the shelves without disconnecting any wires.

In the receiving section of the control center, all transmission circuits except the detectors are shielded. A large terminal block in the console provides readily accessible terminals to which the installer attaches the many leads for control of local A and B equalizers and other leads associated with the receiving line connecting circuits and the loop-back units. A large lead connecting the console framework to the building ground guards against accidental shock to operating personnel.

The loop-back unit is assembled on three fabricated shelves attached to unit mounting bars. The over-all assembly is arranged for mounting on a standard 23-inch relay rack and is located in the control connecting bay of an L-4 bay lineup. The various control circuits are assembled on printed wiring boards, each of which is generally associated with one function such as control of line selection. Interconnections among the boards are made by means of local cables. Partial access is provided to the components on a shelf when it is secured in place. For complete access, mounting screws must be removed and the shelf must be withdrawn. Adequate loops in the local cables permit the withdrawal without disconnecting any leads.

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Solid State Devices

By NORMAN J. CHAPLIN, GRAYDON A. DODSON,
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This paper describes the electrical and physical characteristics of the solid-state devices developed for the L-4 system. These include new high frequency, planar, epitaxial, silicon transistors with low noise figures and intermodulation distortion for use in the various amplifiers. New diodes for power supply, surge protection, and modulation use were also developed.

I. INTRODUCTION

New solid state devices were developed to meet the stringent requirements of the L-4 system. Since the intermodulation distortion and noise are additive over the thousands of miles of system, low noise figure and very low intermodulation distortion limits are placed on the transistors. A low capacitance surge protection diode was developed to provide secondary surge protection for the repeater input. A multiple low capacitance diode was developed as an 18 MHz modulator in the terminal; four diodes were developed for the terminal.

II. TRANSISTORS

Concurrent with the conception and design of the L-4 system, major advances were achieved in transistor technology which have been exploited to reduce system noise, to reduce distortion, and to improve system reliability.

These advances resulted in the development of two families of high frequency, planar, epitaxial, NPN silicon transistors for use in the various amplifiers of the L-4 system.

2.1 *Description*

Two families of transistors were developed to meet the requirements of various L-4 system applications. The first family consists of the 45-types, which were designed for low power, low noise, and low distortion. The second family includes the 46-type transistors, which were designed for medium power, low distortion applications.

The transistor wafers are encapsulated in metallic ceramic packages which were designed for these devices in order to minimize parasitic capacitance and lead inductance, and to provide low thermal impedances while electrically insulating the transistor from the heat sink. The 45-type transistors are encapsulated in an aluminum oxide ceramic-kovar package designated the JETEC TO-112 and the 46-types are encapsulated in a similar package except that a beryllium oxide ceramic was substituted for aluminum oxide to achieve a 20°C per watt thermal impedance required for the medium power operation.[1]

2.2 *45-Type Transistors*

The stringent requirements of the L-4 system amplifiers determine the critical transistor parameters such as low noise figure, high gain, high gain-band product, and low distortion. The 45D transistor is an example of the 45 transistor family; it illustrates how the stringent parameters are met.

The 45D transistor, shown in Fig. 1, consists of eight emitter stripes, $15\mu \times 50\mu$ each. The eight emitter stripes are connected in parallel by the aluminum metalization as are the ten base stripes. The design of the device was dictated by the circuit requirements listed in Table 1.

A typical use of the 45D is in the input stage of the basic repeater, in which a low noise figure is required at a bias current of approximately 20 mA. The relatively high bias current is required in order to avoid distortion in this stage. A noise figure of less than 4.0 dB is achieved by a base resistance of less than 25 ohms. This value of base resistance is achieved by using the eight emitter stripes. The gain-bandwidth product of 900 to 1,300 MHz was achieved by making the width of the base approximately 0.5μ; the variation of the gain-bandwidth product with collector current is illustrated in Fig. 2.

Other transistors in the 45 family are similar to 45D but vary in size, in order to vary current handling ability.

2.3 *46-Type Transistors*

The 46-type transistors fulfill the L-4 system requirements for a medium power transistor which has low distortion, high gain-bandwidth product, and tightly controlled dc gain. The 46E is a good example of the 46 family.

The 46E transistor, shown in Fig. 3, consists of 72 emitters, $15\mu \times$

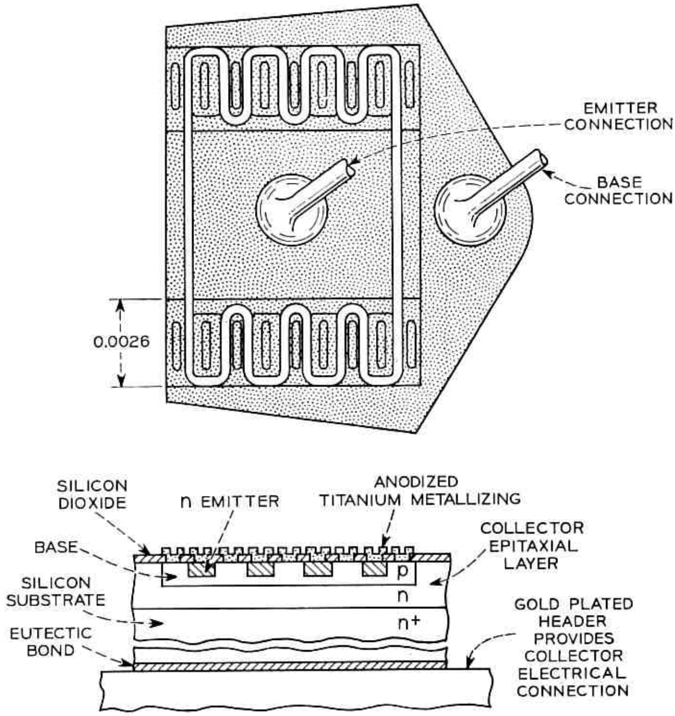

Fig. 1 — The 45D transistor.

15μ, connected in parallel by the aluminum overlay. Base contact is provided by low resistance p-type diffusions under the emitter metalizing instead of metallic stripes as in the 45D. This overlay structure yields a transistor with a large emitter periphery to emitter area ratio and balanced current flow in each of the emitters.

TABLE 1 —TRANSISTOR REQUIREMENTS

Parameter	45D	46E
$V_{CE(SUS)}$	>14 V	>30 V
h_{FE}	100–300	35–75
Noise figure	<4.0 dB	—
f_t	900–1,300 MHz	600–850 MHz
M_{3E} (f = 17.0 MHz)	—	−99 dBm
M_{2E} (f = 17.0 MHz)	—	−53.2 dBm
I_C max	160 Ma	500 mA
$I_C$typical	20 mA	170 mA

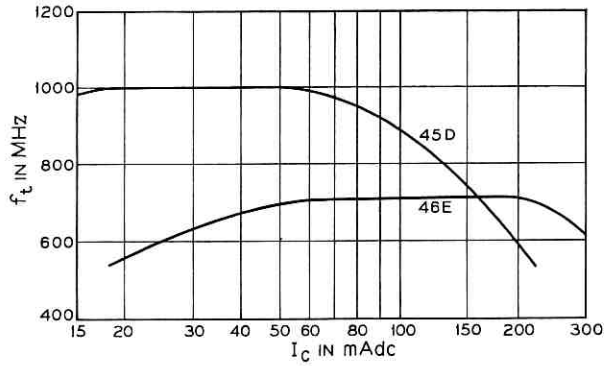

Fig. 2 — Variation of the gain-bandwidth product with collector current.

Important transistor requirements are illustrated in Table I. The dc gain and gain-bandwidth product ranges are required to achieve amplifier gain and phase margins. The stringent third harmonic distortion requirements were met by designing the 46E with a large region of constant gain and gain-bandwidth product as a function of the collector current. This gain-band product characteristic is illustrated in Fig. 2.

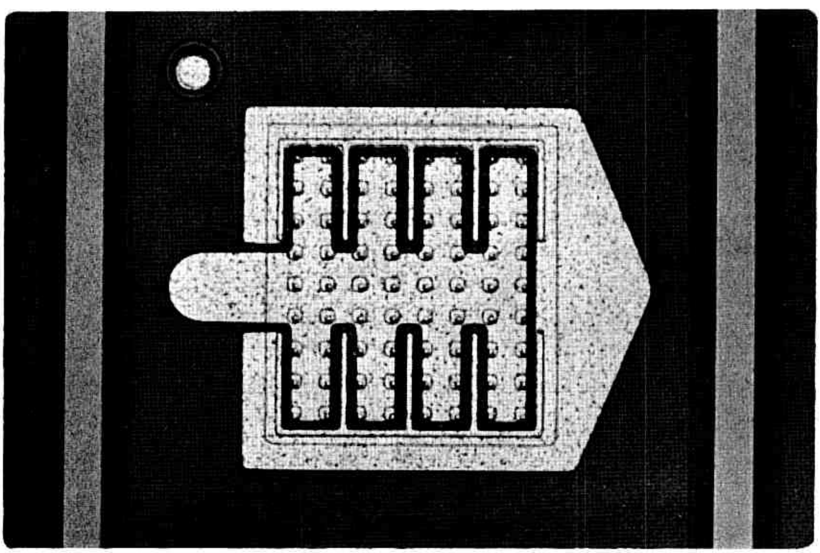

Fig. 3 — The 46E transistor.

2.4 *Transistor Reliability*

During the development of the L-4 transistors several thousand units were subjected to accelerated aging under conditions of both power and temperature stress. Extrapolation of these data predicts a transistor failure rate for both the 45 and 46 types of less than 10 FITS* over a twenty-year period at junction temperatures of 125°C. Reliability information obtained from system use agrees with the extrapolated results.

2.5 *Conclusion*

The 45 and 46 families of transistors have been developed and manufactured to meet the stringent requirements of the L-4 system for low noise, low distortion, and highly reliable transistors.

III. DIODES

Six new diode codes were required for the L-4 carrier system. In general, these break down into three use categories: power supply, surge protection, and modulation. The 460B, 464C, 478B, and 496A diodes were developed for power supply use. The 495A was developed for primary repeater surge protection; the 458E was developed for secondary repeater surge protection. The 460C multiple diode was developed for the mastergroup multiplex terminal (MMX).

3.1 *460 B Diode*

The 460B diode is four high-speed *p-n* junctions connected in a full wave bridge configuration sharing a single TO-55 type encapsulation (shown in Fig. 4). This device is a full wave bridge rectifier in the 20KHz dc-to-dc converters used to power the repeater in the L-4 carrier system.

Design requirements were: breakdown voltage (V_B), > 60 V; output current (I_o), $\leqq 100$ mA ($\leqq 1$ A for lms trouble condition); reverse recovery time, (t_{rr}), $\leqq 100$ ns; saturation current (I_s), $\leqq 15$ μA; and packaged in a four leaded TO-55 package. The breakdown voltage and output current are consistent with the load requirements for the power supply; the reverse recovery time and saturation current are requirements necessary to meet the converter efficiency. The package requirement meets the need for a small easy-to-mount diode package. These requirements were all met by using four gold doped planar epitaxial

* 1 FIT = 1 failure per 10^9 device operating hours.

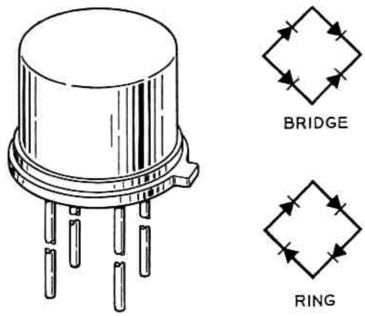

Fig. 4 — Mounting arrangement.

p-n junctions mounted in a metallized ceramic header of a TO-55 package.

3.2 *464 Diode*

The 464C diode is a molded diode assembly consisting of four high-speed high-voltage one-watt diodes (426 Type) connected in a full wave bridge configuration and molded in epoxy. This device serves as a high voltage rectifier for the converter power stages in the terminal station of L-4 carrier system. Design requirements were: breakdown voltage, (V_{BR}), $\geqq$ 800 V; output current (I_o), $\leqq$ 500 mA (650 mA for several hours under trouble conditions); reverse recovery time (t_{rr}), $\leqq$ 100 ns; and packaged into a single molded assembly. The V_B and I_o are consistent with the power supply requirements, whereas t_{rr} was low to improve power supply efficiency.

The molding of completed packages (426 types) served two functions, first to minimize any corona problems and, second, to make an easy to mount package. These requirements were met by molding four 426 type diodes into a four terminal bridge configuration. The structure is shown in Fig. 5.

The silicone resin serves as a high dielectric insulator across the glass seal area of the metal diode package and also mechanically decouples it from the epoxy resin. The epoxy serves as a support for the diode and wiring structure and seals the diode from moisture. An alumina filling agent in the epoxy lowers the thermal impedance of the package. The shell, although not required for the design of the molded assembly, was used to reduce the cost of fabrication. In order to avoid corona, care is exercised during fabrication to avoid pinholes and voids in the molding materials and sharp projections on the metal parts.

3.3 *478B Diode*

The 478B diode is a molded diode assembly consisting of four high-voltage one-watt diodes (426 type) connected in a series string and molded in epoxy. This device goes directly across the output terminals of each power converter to provide a current path for the other series connected converter should that converter's output voltage fail. Design requirements were: breakdown voltage (V_{BR}), $\geqq$ 3,500 V; $I_o \leqq 650$ mA (in trouble condition only for a period of several hours); and molded into a single package assembly. Under normal operating conditions this diode is in the back biased (blocking) condition with no forward current flowing. The requirements were met by molding four 426G diodes into a series string. The diode assembly was fabricated in the same manner as the 464C molded assembly.

3.4 *496A Diode*

The 496A diode consists of two 425L diodes matched for forward voltage drop. This diode is used as a full wave rectifier in the 2.5 KHz dc-to-dc converter to power the terminal bays of the L-4 carrier system. Design requirements were: breakdown voltage (V_{BR}), $\geqq$ 200 V; output current (I_0), $\leqq$ 1.25 A (6A peak current); forward voltage drop (V_f), $\leqq$ 1.2 V at 2.5 A dc; and forward voltage unbalance (V_f (unbal)), $<$ 50 mV at 2.5 A dc. The V_{BR} and I_o requirements are consistent with the power supply requirements and are easily met by the 425L ten watt diodes. The upper limit on V_f is to prevent inefficiencies and power loss in the converter. The forward voltage matching requirement was designed to reduce the second harmonic distortion (5 KHz) and thus further assure its complete removal by filtering. If not completely removed, this signal would appear as noise in the audio band.

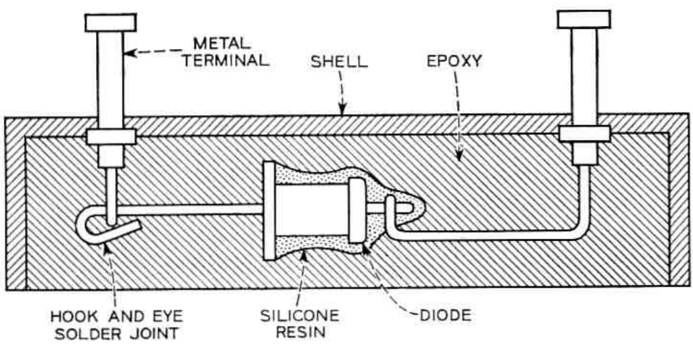

Fig. 5 — Schematic of assembly cross section.

3.5 *495A Diode*

The 495A diode which is used as primary surge protection at the input and output terminals of the L-4 repeaters has a low capacitance. It is a multiple diode consisting of two p-n junctions connected in series with a center tap and encapsulated in a three leaded package. When used at the input of the repeater, the terminals of the 495A diode are wired so that the two p-n junctions are connected parallel-opposing, and act as a bidirectional surge protector. When used at the output of the repeater the center tap of the two p-n junctions is connected to the collector terminal of the output transistor; the other two terminals are appropriately connected to the most positive and to the most negative terminal of the power supply. This configuration thus protects the amplifier by limiting the voltage swing on the collector of the output transistor should a voltage surge be induced in the cable.

Design requirements for this diode were: breakdown voltage (V_{BR}), $\geqq$ 50 V; forward voltage drop (V_F), $\leqq$ 1.0 V at 100 mA dc; saturation current (I_s), $\leqq$ 200 nA dc; capacitance (C), $\leqq$ 12 pf; and nonrepetitive forward current surge of 8 A for 500 μs and 25 A for 10 μs. The breakdown voltage, forward voltage drop, and saturation current are consistent with the diode requirements for this end use. The capacitance requirement was necessary to minimize any attentuation in the transmission path. The nonrepetitive current requirement was necessary to assure that the diode would effectively absorb the surge currents which could be transmitted down the cable. These design requirements were met by using two planar p-n junctions in a three leaded TO-18 package. Figure 6 shows the device outline.

3.6 *485E Diode*

The 458E diode, which has a low capacitance, is used as a secondary surge protector at the input to the L-4 repeater amplifiers. Two 458E diodes are used, connected parallel-opposing, to act as a bidirectional surge protector. These units are connected across the emitter base terminals of the input transistor in the solid state amplifiers. Design requirements were: breakdown voltage (V_{BR}), $\geqq$ 50 V; forward voltage drop (V_F), $\leqq$ 1.0 V at 100 mA; saturation current (I_s), $\leqq$ 50 nA; capacitance (C_o), $\leqq$ 4.0 pF; and nonrepetitive forward voltage (v_f nonrep.) $\leqq$ 2.10 V at 1.0 A peak, t_p = 30 ns maximum. The breakdown voltage, forward voltage drop, and saturation current are consistent with the diode requirements for this end use. The low capacitance requirement was necessary to avoid introducing any attenuation in the transmission path. The nonrepetitive forward voltage requirement was necessary to assure

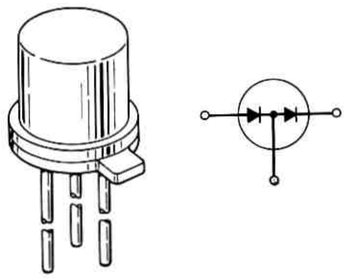

Fig. 6 — The device outline for the 459A diode. All dimensions are in inches.

that the maximum voltage at the input terminals of the solid state amplifier would not exceed one volt (either polarity). All of these requirements were met by using an epitaxial wafer mounted in the small glass diode, package (see Fig. 7).

3.7 *460C Diode*

The 460C diode is a multiple of four p-n junction diodes used as an 18 MHz modulator in the mastergroup multiplex terminal in L-4. The four closely matched p-n junction diodes are connected in a ring configuration and mounted in a four-leaded TO-55 package. Design requirements were: saturation current (I_s), $\leqq$ 20 nA dc at 20 V; capacitance (C_o), < 4.0 pF; reverse recovery time (t_{rr}), < 4.0 n.s. The saturation current requirement is consistent with diode requirements for this end use. The low capacitance and reverse recovery time requirements are necessary to assure that no imbalance in these parameters will result in leakage of the carrier frequency into the output of the modulator. Because the four p-n junctions are wired into a ring configuration these three parameters are measured on the individual diodes before they are mounted in the package.

The carrier leak test is a final test made on the finished 460C diode to

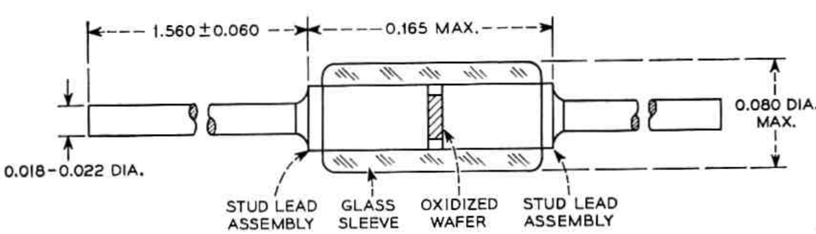

Fig. 7 — Mounting and wiring arrangement.

assure that all parameters, including forward voltage drop, are in balance and the device is functioning properly. These requirements were all met by using four gold doped planar epitaxial p-n junctions mounted in a TO-55 package.

3.8 *Diode Reliability*

Accelerated aging of L-4 system diodes during system development has indicated junction temperature to be the primary failure stress for diodes used within their maximum rated voltage. As indicated in the preceding sections, many of the L-4 diodes are used as protectors during either surge or trouble conditions and hence are called into operation for relatively brief and infrequent intervals of system operation during which they may be safely operated near or above their nominal maximum rated power and temperature.

Diodes not so used are operated at relatively low power levels or are located at system terminals where adequate cooling is available. Information from system use indicates satisfactory system reliability.

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Terminal Arrangements

By W. G. ALBERT, J. B. EVANS, Jr., T. J. HALEY, T. B. MERRICK, and T. H. SIMMONDS, Jr.

(Manuscript received July 25, 1968)

New terminal equipment has been developed to provide the six-mastergroup signal that the L-4 repeatered line is capable of carrying. A major component of this equipment is the MMX-2 mastergroup multiplex, which translates six basic 600-channel mastergroups to their line frequency assignments in the range 0.564 to 17.548 MHz. Another component is the line connecting equipment, which contributes to shaping the line signal frequency characteristic and provides for blocking mastergroups. The purpose of this blocking is to provide dropping and branching with recovery of the spectrum for transmission of other mastergroups.

Mastergroup connectors are available for flexible interconnection of basic mastergroups without demodulation to supergroups. Finally, existing line protection switching equipment has been modified for use with the L-4 Coaxial System.

I. INTRODUCTION

The L-4 coaxial line is described in Ref. 1 as a transmission facility with a message capacity of six mastergroups. These six mastergroups are placed on the line in the frequency assignment shown in Fig. 1. Each mastergroup starts as a U600 basic mastergroup, formed by L-type multiplex (LMX) frequency-division multiplex equipment. An important function of the L-4 terminal equipment is the translation to line-frequency assignment of the six mastergroups. Earlier mastergroup multiplexing equipment, designed for the L-3 coaxial cable system and later used for TH microwave radio systems, has a capacity of three mastergroups.

The U600 basic mastergroup consists of 600 single-sideband channels stacked up in the frequency range 564-3084 kHz in 12-channel groups and 60-channel supergroups by frequency division multiplex techniques, using LMX equipment.[2] This frequency assignment is now

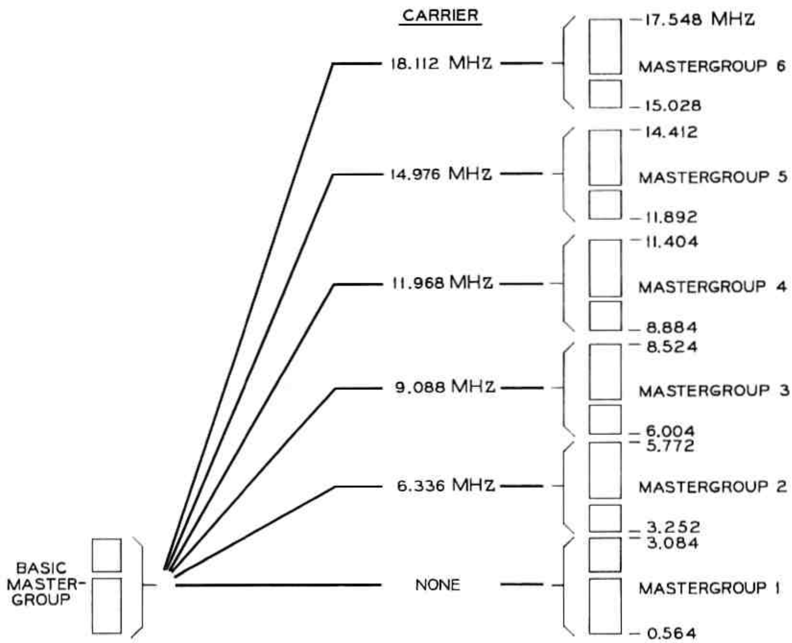

Fig. 1 — Six-mastergroup line frequency spectrum.

common to L-3 and L-4 coaxial systems, TH and TD-3 radio systems, and some TD-2 radio systems. This permits the greatest part of the terminal station equipment to be relatively high demand, common use equipment.

To provide the six-mastergroup multiplexing function, a transistor mastergroup frequency-division multiplex has been developed and manufactured. It is known as MMX-2 (mastergroup multiplex-2) and is distinguished from the earlier, electron tube equipment by naming the latter retroactively MMX-1.

MMX-2 differs from MMX-1 not only in number of mastergroups, but also in line frequency spectrum. Mastergroup 1 is common to all multimastergroup systems,* and is in fact identical in frequency spectrum with the basic mastergroup; that is, it is applied to the line without frequency translation. The frequency spacing between mastergroups, a constant 80 kHz in MMX-1-equipped systems, has been increased to a constant four percent (approximately) for L-4 coaxial

* This is not strictly true, since the earlier systems add a supergroup below 564 kHz, and define the resulting 660-channel signal as mastergroup 1.

and other MMX-2-equipped systems.* This was done to permit blocking and branching of mastergroups at line frequencies with practically realizable filters.

Thus two major functions of the L-4 terminal have been described or implied above:

(*i*) Multiplexing of up to six basic 600-channel mastergroups is provided by the MMX-2 equipment. The MMX-2 equipment also provides demultiplexing for the other direction of transmission.

(*ii*) To reduce costs, as well as to improve performance by virtue of reducing and simplifying the equipment in a built-up connection, line blocking and branching of mastergroups is provided by the L-4 line connecting equipment. Thus only dropped mastergroups need be demultiplexed; others go through completely passive filters to continue along a backbone route ("through mastergroups") or go through filters and amplifiers to side-legs or branch routes (branched mastergroups).

The line connecting equipment is physically separate from the mastergroup multiplex and is contained in the line transmitting, receiving, and common bays. In addition to mastergroup blocking and branching, as discussed above, line connecting equipment provides several additional functions:

(*i*) Two of the three full-time line pilots, 11.648 MHz and 20.448 MHz, are generated by precision, free-running crystal oscillators in the common bay and distributed to the transmitting bays.

(*ii*) A 512 kHz pilot, which is used to mutually synchronize MMX and LMX circuits in different offices, for actuating the L-4 terminal low band-edge regulators, and for actuating protection switches in the line connecting circuits, is stabilized in amplitude in the common bay and distributed to the transmitting bays. This pilot is generated by an office primary frequency supply not specifically a part of the L-4 equipment. The three pilots are either bypassed through the office or blocked and reinserted.

(*iii*) Signal shaping, consisting of 6.1 dB square-root-of-frequency pre-emphasis and de-emphasis, is included in line connecting component equipment called MMX-2 connecting circuits.†

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* MMX-2 equipment for microwave radio, providing three-mastergroup capability and differing slightly from the L-4 version, has also been developed.

† Additional shaping in the transmitting main repeater brings the pre-emphasis to about 18 dB as it leaves the office.

(*iv*) Connection of the entire line signal to the receiving remote-control center is provided, as well as connection of the command channels to and from the remote-control system.[3]

(*v*) Test and patching access is provided.

(*vi*) Equalization for the slope in the transmission frequency characteristic of interbay office cabling is provided.

(*vii*) Access for emergency restoration of service of L-4 systems over other L-4 systems (baseband restoration, a term carried over from radio systems) is provided. Restoration of other systems, and of L-4 over other systems, is available also. This involves restacking of mastergroups, however, and is provided by MMX equipment.

Two additional important functions are provided by terminal equipment not specifically line connecting or multiplex:

(*i*) Interconnection with L-multiplex circuits has already been mentioned. Interconnection of basic mastergroups on the office side of the MMX circuits can be made via mastergroup connectors between L-4 coaxial systems and radio systems or other coaxial systems. This provides a means of routing mastergroups without using LMX equipment, which produces cost savings and performance improvements. It competes favorably with line branching because it is more flexible in administration (reassignments involve virtually only office cabling changes) and at present represents the only means of interconnecting intact mastergroups between L-4 coaxial and other systems.

(*ii*) Modified L-3 electron tube equipment, in conjunction with the line connecting circuits, provides automatic standby line protection of L-4 repeatered lines for up to nine working coaxial pairs (a fully equipped 20-coax cable).

Figures 2 and 3 show the L-4 terminal arrangements as described above, including MMX, line connecting, protection line switching, and mastergroup connector functions.

II. MMX-2 MASTERGROUP MULTIPLEX

The MMX-2 mastergroup multiplex provides the steps of modulation and demodulation between the basic mastergroup and the L-4 line signal. In addition to modulators and demodulators, it provides redundant, automatically switched carrier and pilot supplies and automatic transmission switching circuits.

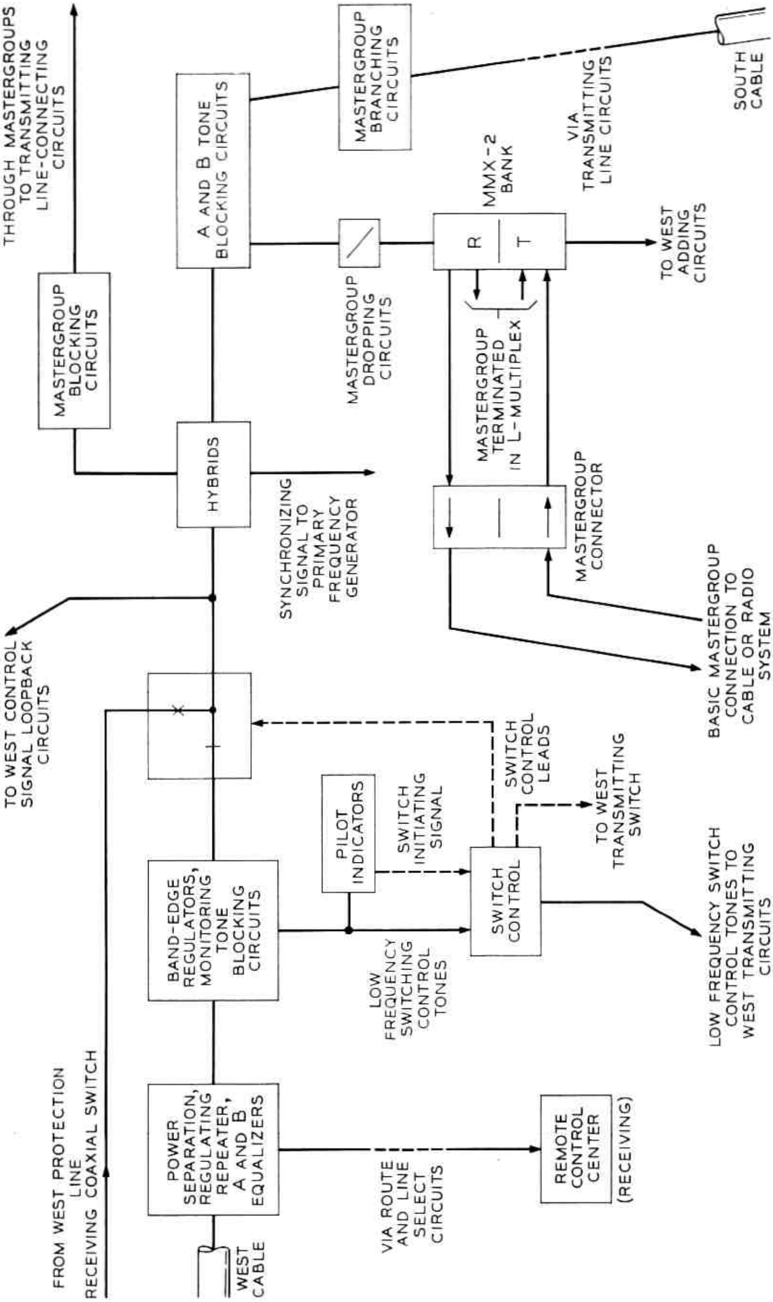

Fig. 2 — Receiving terminal.

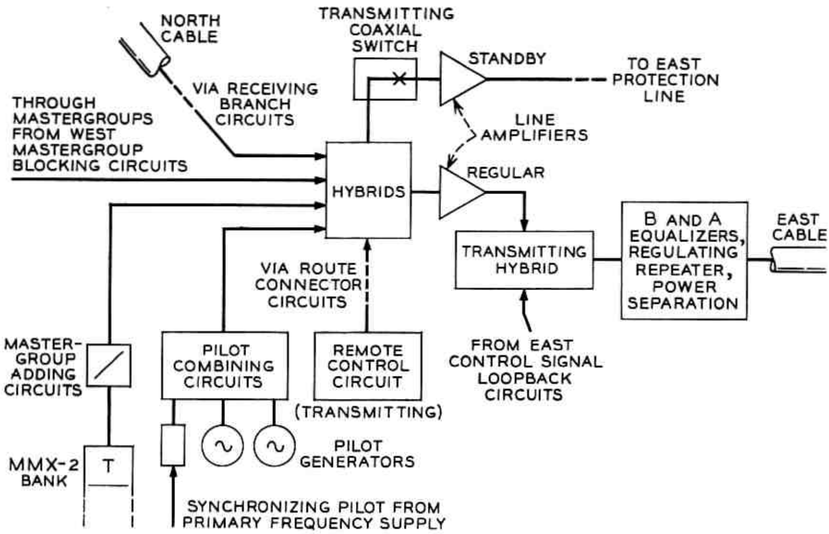

Fig. 3 — Transmitting terminal.

2.1 *Transmission Circuits*

2.1.1 *Transmitting Circuits*

The MMX-2 transmitting circuits are shown in the block diagram, Fig. 4. The output of the LMX equipment, a basic mastergroup, is connected into one input of a hybrid after having the mastergroup pilot (2.840 MHz) bridged on. The second input of this hybrid is available for emergency patching. The net side of the hybrid is connected to a test jack through a 10.8 dB pad providing a −35 dB transmission level test point for measuring the input signal. The primary output of the hybrid is fed into a second hybrid which provides two outputs. One output feeds into the regular modulator circuit, and the other feeds into the spare modulator circuit via the transmission switching circuit.

The modulator circuit consists of a low-pass filter, a diode modulator, a low level amplifier, a band-pass filter, and a high level amplifier, in that order, except for mastergroup 1. Mastergroup 1 uses only a low-pass filter and an amplifier.

The output of the modulator circuit is applied to the pilot hybrid. The hybrid provides two outputs. One output is connected to the transmission switching circuit and from there to the transmitting

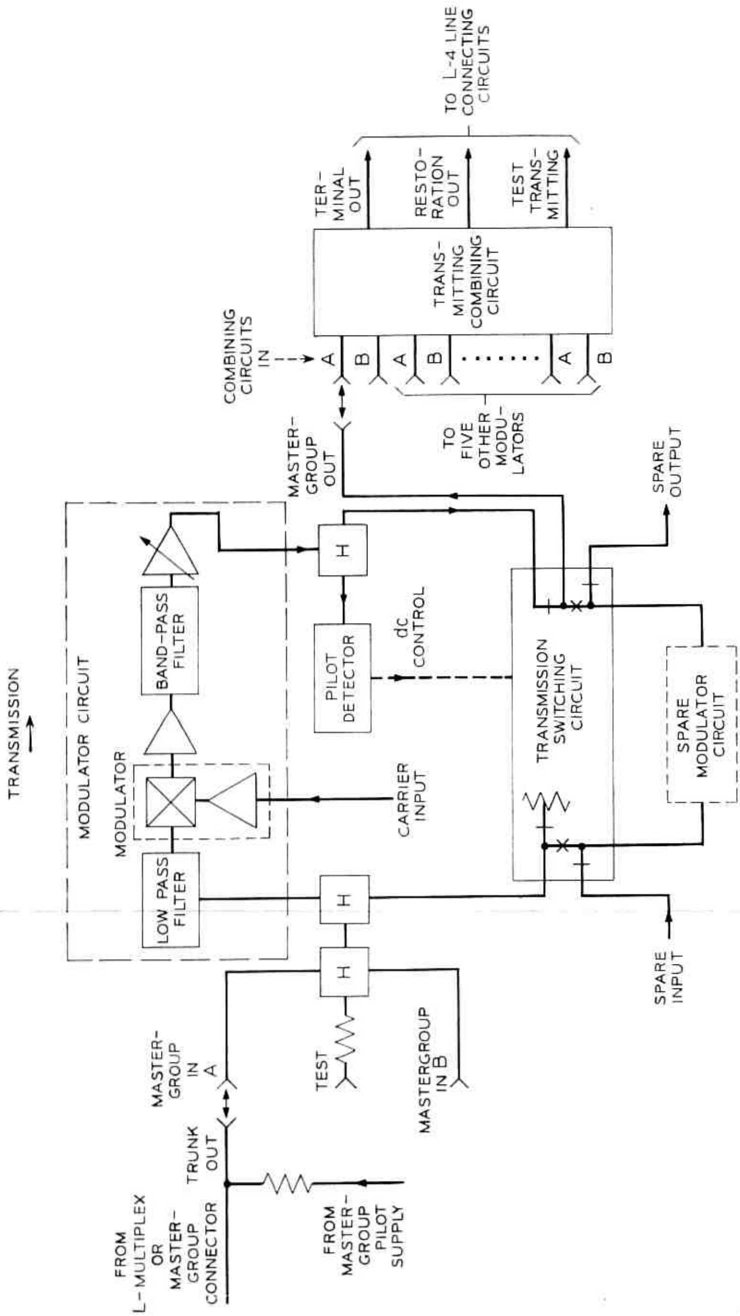

Fig. 4 — MMX-2 transmitting circuits.

combining circuit. The second output feeds the pilot detector which controls the operation of the transmission switching circuit.

The combining circuit provides twelve input ports and three output ports. Six of the input ports are normally connected to modulator circuits operating at the six different mastergroup frequencies. The remaining six input ports are normally terminated and are available for emergency patching. Since the frequency responses of the input ports vary slightly, they have been matched to the six mastergroup frequencies to optimize the overall circuit performance. The primary output port of the circuit is connected to the L-4 line via the line connecting equipment. A second port provides a test output point at −45 dB transmission level. The third output port provides for permanent connection to service restoration equipment.

2.1.2 *Receiving Circuits*

The broadband signal received from the L-4 line is connected to the receiving splitting circuit. This circuit provides 12 equal-level outputs for one input signal. Six of these outputs are connected to regular demodulator circuits, while the other six are connected to the spare demodulator circuits via the receiving transmission switching circuit.

The demodulator circuit consists of an input amplifier, a band-pass filter, a diode demodulator, a low-pass filter, a low level amplifier, and a regulating amplifier, except for mastergroup 1. For mastergroup 1, the circuit consists only of a band-pass filter, a low level amplifier, and a regulating amplifier, since no frequency translation is required. The output of the regulator is connected via the transmission switching circuit to an output hybrid which provides a normal output that connects to LMX equipment and a spare output which can be used for monitoring the demodulator output as well as for emergency patching.

2.1.3 *Modulator Circuit*

The low-pass filter is provided to suppress spurious signals above the basic mastergroup frequency band which may be generated in preceding transmission circuits. It is a constant resistance design and so maintains a good impedance, greater than 26 dB return loss, at frequencies up to 35 MHz with the exception of a band between approximately 4.2 and 4.8 MHz, where it falls to a minimum of 15 dB return loss. This provides a good resistive termination for the following modulator without necessitating the addition of large amounts of padding which would be required otherwise. By maintaining a good impedance up to and beyond 35 MHz, the filter properly terminates not only the wanted but the unwanted sidebands produced by the modulator.

The modulator is a double balanced ring modulator which uses a 460C multiple diode. The diode consists of four epitaxial silicon, planar wafers, connected in a ring and encapsulated in a TO-55 enclosure. The diode elements exhibit a reverse recovery time (t_{rr}) of less than 4 nanoseconds and capacitance at 0 volts of less than 4 picofarads. Broadband transformers are used at both the input and output of the modulator, which allows the use of identical modulators for all of the mastergroups. The modulator has a conversion loss of approximately 5.1 dB* at any of the mastergroup frequencies, with a variation of ± 0.05 dB maximum across the mastergroup frequency band. It maintains a carrier balance of greater than 38 dB at the highest carrier frequency (18.112 MHz), and greater than 45 dB at the lowest (6.336 MHz).

The modulator package also includes a carrier amplifier which provides approximately 17 dB of gain. This amplifier serves two purposes. First, it reduces the required level of carrier signal which must be cabled from the carrier supply to the modulator, with a consequent reduction in the amount of carrier signal coupled into low level transmission cables within the MMX-2. Second, it provides isolation in the carrier path between modulators fed from a common carrier distribution circuit. The amplifier consists of two common emitter stages with local feedback. It operates with a +3 dBm input signal to drive the modulator at +20 dBm. At this level, the modulator exhibits a stiffness of greater than 10:1 with changes in carrier level.

A set of useful modulation coefficients for the modulator can be defined as the ratio of unwanted to wanted sideband powers, with the wanted sideband power at 0 dBm at the output of the modulator. That is

$$M_{2(MOD)} = \frac{P_{(F-2A)}}{P_{(F-A)}}, \qquad M_{3(MOD)} = \frac{P_{(F-3A)}}{P_{(F-A)}}$$

where F is the carrier frequency and A is the input signal frequency. These coefficients were measured and found to follow the expected dB-per-dB change for $M_{2(MOD)}$ and 2 dB-per-dB change for $M_{3(MOD)}$ with changes in input signal level. In addition, close agreement was found between measured and calculated $M_{A-B(MOD)}$, $M_{2A-B(MOD)}$, and $M_{A+B-C(MOD)}$, where the calculation was based upon measured $M_{2(MOD)}$, $M_{3(MOD)}$, and a power series expansion. For a typical modulator, $M_{2(MOD)}$ was found to vary by approximately 5 dB with carrier frequency and $M_{3(MOD)}$ by approximately 8 dB. With the worst case

* Theoretical minimum conversion loss is 3.9 dB using resistive terminations.

carrier applied, $M_{2(MOD)}$ is typically -45 dB and $M_{3(MOD)}$ typically -42 dB. MMX-2 noise computed using these parameters agrees to within 3 dB of noise measured using noise loading. With noise loading, the back-to-back MMX-2 noise was found to be between 11 and 15 dBrnC0, depending on the mastergroup measured, which meets the terminal objective of 19 dBrnC0 with adequate margin.

In order to meet the MMX-2 noise requirement, it was necessary to operate the modulator at a relatively low signal level to limit its modulation noise contribution. To achieve this and yet avoid introducing excessive thermal noise, it was decided to follow the modulator with a low level amplifier before filtering. Two requirements are placed on the low level amplifier. It must have (*i*) a relatively low noise figure, and (*ii*) good modulation characteristics, since it must carry the full modulator output, including the carrier leak and the unwanted sidebands.

The 266 type amplifier was developed for this use, as well as for use in several other positions in the MMX-2 circuitry. The basic amplifier has a noise figure of 10 dB and modulation characteristics such that $M_{3(MOD)}$ is better than -93 dB at 17 MHz and $M_{2(MOD)}$ is better than -80 dB at 0.7 MHz. The amplifier was constructed such that padding could be added at either or both the input and output if required. For the low level amplifier, it was found that 5 dB of padding was required at the input to provide a proper termination of the higher order modulator sidebands. The amplifier, coded 266A, consequently has a noise figure of 15 dB and input return loss better than 35 dB across the L-4 frequency band. A second 266 amplifier, the 266C, is used as the high level output amplifier in the modulator circuit. This amplifier is adjustable over ± 2 dB around a nominal 12dB gain.

The band-pass filters used in the modulator circuits are the only components which differ from mastergroup to mastergroup. They are, however, similar in their requirements stated in terms of the modulator output signals. These are:

Type of product	Minimum discrimination (dB)
Signal leak (S)	60
Carrier leak (C)	70
Unwanted sideband ($C + S$)	84
$2C$	70
$2C + S$	44
$3C + S$	74
$3C$	85
$4C$	65

In addition, a requirement is imposed by the use of the filters in the demodulators as well as the modulators; that is, that the band of signals in the adjacent mastergroup be suppressed by 84 dB minimum (see Section 2.5).

2.1.4 *Demodulator Circuit*

The demodulator circuit uses many of the component parts of the modulator circuit. The diode demodulator is identical to the modulator except for a slight mechanical rearrangement. The amplifiers are 266 types which differ from those used in the modulator circuit only by the amount of input and output padding required in their specific circuit application.

The unique feature of the demodulator is the transmission regulator. This is the first use of a transmission regulator for the mastergroup frequency band.

2.1.5 *Transmission Regulators*

The regulating circuitry serves two functions. It provides flat gain regulation and it develops the pilot out-of-range signal to activate the receiving transmission switches and alarms. Figure 5 is a diagram of the regulator.

The feedback regulator uses the demodulated mastergroup pilot at 2.840 MHz to control the output level of each mastergroup from the receiving side. The regulator circuit is conventional and uses a pilot pick-off-pad, pilot selection filter, pilot amplifier, detector, and comparison circuit. The comparison circuit controls a current to a thermistor in the feedback circuit of the transmission amplifier.

2.1.5.1 *Regulation Control Loop.* The pilot pick-off-pad provides a termination at the output of the high impedance transmission amplifier. The pad loss is 15 dB, which gives 30 dB return loss against impedance variations of the crystal pilot filter. The pilot amplifier has two sections separated by a series tuned circuit. Each section has negative feedback to stabilize the gain, which is approximately 50 dB for the two sections when measured with a 75-ohm termination. A rectifier connected to the high impedance output of the pilot amplifier converts the ac signal to dc in the adjustable load resistance.

The output impedance of the pilot amplifier is made high by series feedback; the detected direct current is practically independent of the load impedance and the temperature-sensitive rectifying diode drop. The dc voltage, however, is dependent on the load resistance, which is made variable to obtain the correct regulator output power. The detector and comparison circuit reference is the battery return. This

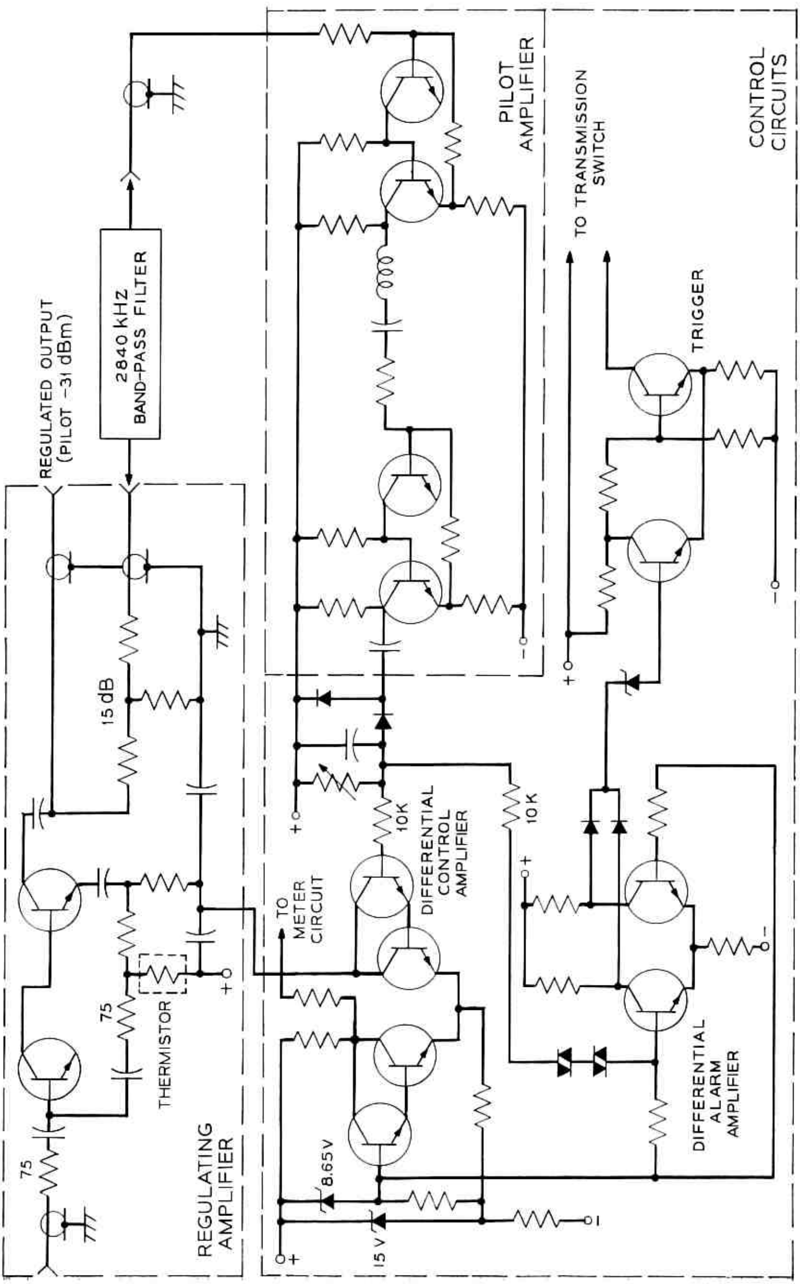

Fig. 5 — Mastergroup transmission regulator circuit.

is done to eliminate ground loop problems. The comparison circuit input is fed from the rectifier through a 10,000-ohm isolation resistor; the comparison circuit consists of a pair of Darlington connected transistors in a differential amplifier configuration. A stable reference voltage from a low temperature coefficient breakdown diode is fed to the second input of the differential amplifier.

The reference side of the differential amplifier provides a current to the meter in the pilot and carrier measuring circuit which gives an up-scale indication for a high pilot input voltage. The control current path is from the other side of the differential amplifier to the thermistor in the regulating amplifier.

2.1.5.2 *Regulating Amplifier.* The regulating amplifier uses series feedback on the second stage and shunt feedback on the first and is similar to the transmission (266 type) amplifiers used throughout the terminal equipment. The maximum output power required corresponds to a +15 dBm sine wave, while the average message output power is near 0 dBm.

This relatively high output power precluded placing the regulating thermistor element directly in the path of the output ac current at the second stage emitter. The sensitive thermistor element used would have heated and allowed the total mastergroup power to affect the regulator gain. Instead, the feedback network was changed from the basic L Form used in the 266 type amplifiers to the configuration shown in Fig. 5. This allows the control current to be fed in at a zero signal voltage point while the maximum thermistor signal voltage is no greater than the input voltage.

The regulator can pass a +20 dBm signal and has modulation coefficients of $M_{2E} = -75$ dB and $M_{3E} = -95$ dB. The stiffness is approximately 30 : 1 over a ± 3 dB range of input, and the total gain variation is ± 4 dB.

2.1.5.3 *Regulator Alarm Function.* The out-of-range alarm signal is derived by comparison of the detector output and reference voltage in another differential amplifier. The collectors of this differential amplifier are connected by a diode OR circuit. The higher of these two voltages is used to set a Schmitt trigger which opens an alarm relay. The alarm is designed to trigger at a pilot output error of ± 3 dB.

2.1.6 *Transmission Switching Circuits*

The transmission switching circuits in the MMX-2 represent a departure in terminal design. While protection switching is not new to the Bell System, switching in multiplex transmission circuits is new. Prior to the MMX-2, automatic protection switching in terminal circuits was confined to carrier and pilot supplies.

The transmission switching has one spare for each three regular circuits. It operates in a modulator-by-modulator or demodulator-by-demodulator mode rather than sparing a full mastergroup bank. The main features and method of operation are basically the same for transmitting and receiving circuit switching.

2.1.6.1 *Transmission Switching Control.* The transmission switching circuits operate in both an automatic and a manual mode. In the automatic mode, the transmitting switching circuit is controlled by the pilot detector circuit, while the receiving switching circuit is controlled by the regulator, which is described in Section 2.1.5.

The pilot detector consists of a demodulator, a 2.840 MHz pilot pick-off filter, an amplifier, and a rectifier-Schmitt trigger circuit. A line frequency output from the pilot hybrid is fed into the demodulator, which translates the signal back into the basic mastergroup frequency band. The demodulator is the same type of circuit described in Section 2.1.4. The 2.840 MHz pilot is selected by the crystal pick-off filter and amplified. The amplifier feeds an 864 network, which consists of a rectifier, a Schmitt trigger, and a relay driver.

Under normal operating conditions, the input stage of the Schmitt trigger is held in conduction by the rectifier. A drop in input pilot level by more than 5.5 dB will cause the Schmitt trigger circuit to change state. With the trigger circuit in its normal state, the relay driver, which has the transmission switching control relay as its collector load, is biased on, operating the relay. When the trigger circuit reverses, the driver releases the control relay. When the input signal rises to within 0.5 dB of its normal level, the trigger circuit reverts to its normal state, reoperating the control relay.

The switching is controlled manually by operating a three-position switch physically associated with the regular modulator. The switch positions are designated LOCK, NORM, and MAN TR. In the LOCK position, the transmission switching is disabled. The NORM position is the regular mode of operation; automatic switching is enabled. The MAN TR position switches the service from the regular to the spare modulator. When the switch is thrown into the MAN TR position, it provides a closure which simulates a release of the control relay; the switching is then accomplished in the same manner as for an automatic switch.

2.1.6.2 *Switching Sequence.* when a transfer is made, the signal is connected through the spare modulator to the output via five relays. The final relay is controlled by a series connection of contacts from the first four. This insures that the signal is connected into the final relay before it is operated. The final relay carries transmission through early-make-break contacts and, consequently, the signals through the regular and spare modulators are double fed to the output and simultaneously double terminated. This condition is removed only when the break contacts have operated. By maintaining as nearly as possible identical amplitude and phase through the two transmission paths, transmission disturbance is minimized during operation of the switching. This type of operation is highly desirable in the manual switching mode used to free a regular modulator circuit for maintenance. Obviously, it is of no real value on operation in the automatic mode, since this implies that transmission has already been lost. Since the same sequence occurs when the switch is released either automatically or manually, it minimizes the hit in both modes upon return to normal.

2.1.6.3 *Switching Alarms and Indicators.* The alarms are arranged to give a minor alarm for any failure which does not cause a loss of service. For example, the failure of a regular modulator with a successful switch to the spare circuit would initiate a minor alarm. A failure which involves loss of service brings in a major alarm. That is, if a switch did not occur upon failure, or did occur but returned to normal because of subsequent failure of the spare, a major alarm would be initiated. Alarm lamps are associated with each modulator as an aid to trouble location; a common bay alarm lamp is also provided.

In addition to the alarm lamps, several other indications are provided. Indicator lamps associated with the manual switch control light when the switch is in other than the NORM position. A CLEAR lamp is included which lights when a failure has occurred and then has restored to normal. The lamp remains lighted until manually released. It is intended as an aid in locating circuits which develop intermittent troubles which clear before a maintenance man has the opportunity to observe alarm lamps. A TR lamp lights when the switching is operated.

2.2 *Carrier and Pilot Supply*

The MMX-2 carrier supply is composed of two major parts (see Fig. 6). The first is called the binary generator. Its function is to generate all the binary multiples (2^0, 2^1, 2^2, . . .) of its 64 kHz input

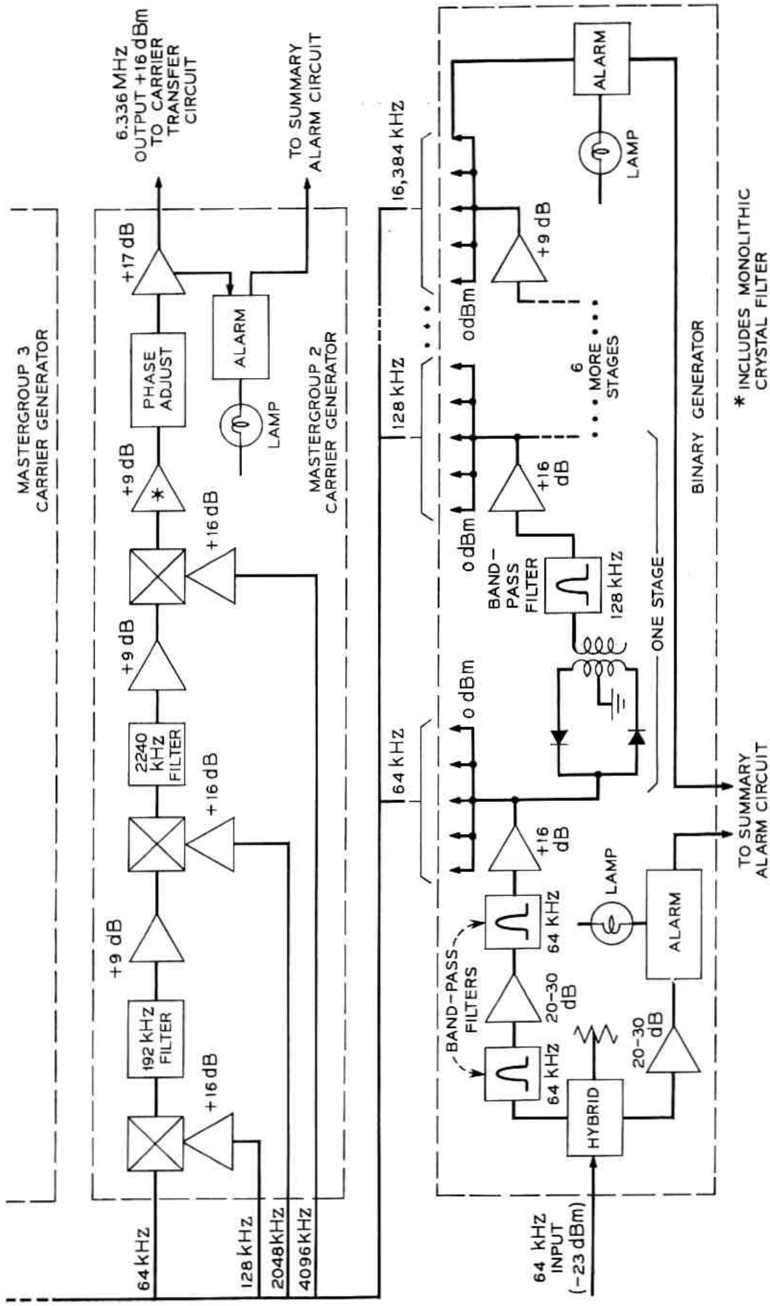

Fig. 6 — Carrier supply circuits.

through 16,384 kHz ($2^8 \times 64$ kHz) and distribute certain of them to the individual carrier generators. The second part consists of five carrier generators, one for each carrier frequency required. These combine three or four input frequencies in modulators to obtain the mastergroup carrier frequencies.

2.2.1 *Binary Generator*

The binary generator is made up of eight similar stages. Each stage filters, amplifies, distributes and rectifies the frequency from the previous stage. The second harmonic from the rectification process becomes the input frequency to the succeeding stage. The filters required are described in Section 2.5.3. The amplifiers are required to have about 16 dB of gain. Each has two stages, the first with local shunt and the second with adjustable local series negative feedback. The amplifiers are identical for all stages requiring rectification, except for transformers.

There are two outputs taken from each amplifier, one at +8 dBm at 15 ohms and the other at +15 dBm at 75 ohms. The first is fed to a series tuned distribution circuit with five outputs, one of which is used for a test jack. The other outputs are connected to the carrier generators. The other amplifier output is fed to the full wave rectifier.

The rectification and filtering provides the second harmonic at about 0 dBm for the succeeding stage. The signal level is such that the rectifying diodes actually overload slightly in that they are heavily saturated; one does not turn off before the other conducts on the opposite half cycle. Significant compression results with about 0.8 dB change in output for a 1.0 dB change in input. Thus input level variations are successively reduced in each stage. This is a desirable level of operation, since a somewhat lower level would have resulted in expansion of input variations.

The first stage of the binary generator actually requires more amplification and filtering than the others, since the input of 64 kHz from the primary frequency supply is at a relatively low level (−23 dBm at 135 ohms). This distribution level was chosen as an optimum compromise between a high level which could cause interference in other circuits and a low level which would be subject to noise pick-up. The preamplifier is adapted from previous designs in N carrier. It is a very useful design which uses two stages with hybrid feedback around both. Its gain is variable with a range of approximately 9 dB.

The extra input signal filtering is provided by two crystal filters to ensure that no spurious signals are accepted. In addition, the power leads to the preamplifier and first three multipliers are specially filtered, since the large multiplication ratio makes them especially susceptible to phase jitter pick-up. Measurements show that multiplication of input jitter is the dominant source of phase jitter in the MMX-2 carrier supply. The last stage, at 16,384 kHz, has no rectifier and only amplifies and distributes the filtered signal from the previous stage. This 9 dB amplifier is basically a utility design developed for use throughout the carrier supply. A single transistor with both series and adjustable shunt feedback is used. Computer calculations were used to find the optimum values of shunt and series feedback elements to be used to obtain the required gain range at maximum return loss.

2.2.2 *Carrier Generators*

The carrier generators for each mastergroup frequency are very similar. As an example, the mastergroup 6 carrier at 18,112 kHz (the 283rd harmonic of 64 kHz) is formed as follows, using double-balanced ring modulators. First, frequencies of 64 and 256 kHz from the binary generator are mixed, and the sum frequency at 320 kHz is selected by a band-pass filter. This 320 kHz is mixed with 2048 kHz from the binary generator to obtain 1728 kHz, which is mixed with 16,384 kHz from the binary generator to obtain the 18,112 kHz carrier frequency. The mastergroup 3 carrier generator requires only three input frequencies; all others require four. Each modulator unit contains a ring modulator with two transformers and a fixed gain (16 dB) carrier amplifier very similar to the one used in the binary generator. The input power levels are 0.0 dBm for both signal and carrier inputs. The modulator (8 dB) and filter losses (0.5 to 3 dB) are made up by adjustable (9 ± 3 dB) single stage amplifiers (262 type) similar in design to the 9 dB amplifier described in Section 2.2.1. The electrical filters used between modulators, as well as the monolithic crystal filter in the amplifier following the final modulator, are described in Section 2.5.3.

The final amplifier (261A) in the carrier generator provides the 13 to 18 dB gain required to drive the distribution circuits and the alarm circuits.

2.2.3 *Carrier Transfer and Alarm Circuits*

The MMX-2 carrier supply is composed of two complete binary generators, each serving one set of carrier generators. This is done to improve the reliability and to allow replacement of portions of one carrier supply while the other maintains service. A carrier transfer unit connects either the A or B carrier generator to the plug-in distribution circuit (see Fig. 7). This distribution circuit, contained in the carrier transfer unit, provides a total of 15 outputs from a three-way hybrid circuit. Test taps are provided on both the distribution circuit and idle carrier generator outputs.

Several alarm circuits are provided as part of the carrier supply. Each binary generator has an input and output alarm. Each carrier generator has an output alarm, and one tap of the distribution circuit is used to generate a major carrier alarm. Each alarm has an associated alarm lamp, which is connected to the alarm circuits so as to cause only the alarm lamp in the first affected unit to light, thereby providing a definite indication of the trouble location. The carrier generator alarms drive the carrier transfer circuit to select the working carrier generator if one fails. IDLE lamps indicate which units are not in service. Provision is included for manual transfer and for prevention of manual transfer to a failed carrier generator.

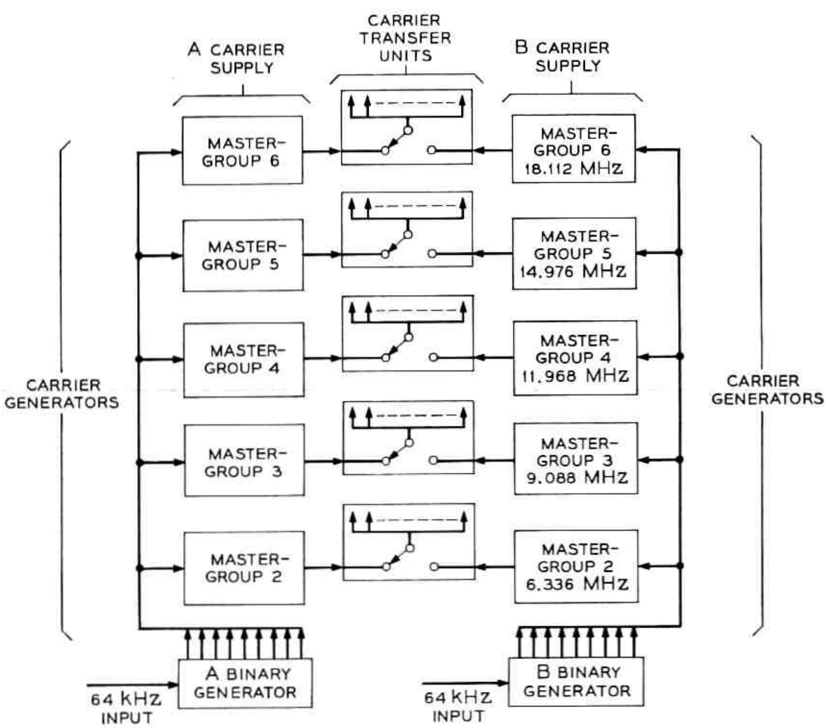

Fig. 7 — Carrier supply arrangement.

Before manual transfer can be made, the carriers must be brought into approximate (±20°) phase coincidence by manual adjustment of phase shifters in the carrier generators. This prevents maintenance transfers from seriously affecting service. A manual inhibit circuit in the carrier transfer unit prohibits transfer until the adjustment is made.

Emergency transfer switches are also provided for use in case transfer must be made and phase coincidence cannot be obtained. The manual inhibit circuit has a meter on the front of the bay to indicate when phase coincidence has been obtained. This meter is also sensitive to level differences between the two sides of the carrier supply. Thus, when a good phase null cannot be obtained because of a level difference, a check of the carrier amplitudes should be made.

In the MMX-2 carrier supply, there is no preferred side (except A when both are failed). Thus when a carrier generator is returned to operation, no automatic transfer is made. Each of the carrier generator alarm circuits is connected to the pilot monitoring panel and to the summary alarm circuit which is described in Section 2.3.

2.2.4 *Mastergroup Pilot Generator*

A mastergroup pilot (2840 kHz, −20 dBm at the 0 dB transmission level point) is used to indicate the level of the entire mastergroup. This pilot is added to the transmitted mastergroup at the input to the transmitting circuits before modulation. Thus it appears in the transmitted line spectrum shifted by the mastergroup carrier frequency. This pilot must have very accurate and stable magnitude (±0.02 dB) and frequency (±7 Hz) as well as high reliability. It is used to activate the mastergroup regulator at the receiving MMX-2 terminal.

Figure 8 shows the mastergroup pilot generator arrangement. The 2840 kHz signal is generated directly in unsynchronized crystal oscillators. The oscillator circuit uses a common emitter oscillator stage with output limiting, followed by a buffer amplifier driving a temperature compensated limiter. An output filter and level control follow the final limiter. The dc power for the oscillator circuit is regulated by a breakdown diode.

Each supply contains two oscillators and associated alarm circuits. These alarm circuits control a transfer circuit in a manner similar to the mastergroup carrier supply transfer control. The actual relay transfer switches required a very careful electrical and mechanical design in order to obtain the required 60 dB isolation. This high degree of isolation is required to prevent the idle oscillator output from

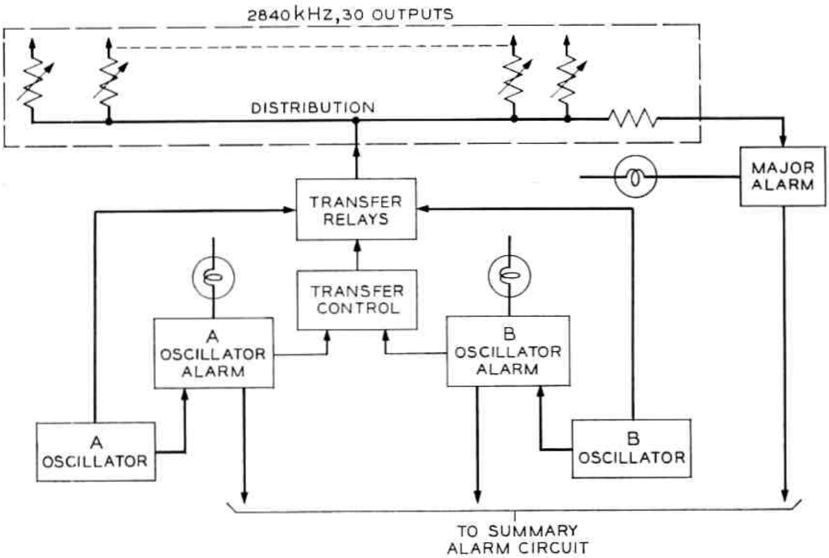

Fig. 8 — Mastergroup pilot generator.

beating with the working oscillator output. The idle oscillator output is fed through the transfer circuit and a pad to a front panel test jack.

The working output from the transfer switch connects to a distribution circuit that has 31 outputs. Twenty-four of these are connected to the transmitting mastergroup trunk inputs. Six others are connected through relays to spare mastergroup trunks. The relays are used to remove local pilot from these trunks if pilot is already present in the input. Each tap in the distribution circuit is individually adjustable. The final tap is used for a loss of pilot (major alarm) indication.

2.3 *Alarms and Maintenance*

The summary alarm circuit (SAC) performs the function of collecting all the electronic alarms in the MMX-2 bay and processing them for connection to external alarm circuits. The external alarm circuits are the office audible, visible, annunciator systems, and the remote alarm system. In addition, some remote status and command signals are connected through a portion of the SAC terminal strips. The only bay alarms which are separated from this unit are the fuse alarms, which are connected through their own relays at the top of the bay.

The electronic alarms in the bay are grouped according to function and classification. These are: transmitting major and minor, receiving major and minor, carrier major and minor, pilot major and minor, and dc to dc converter alarms which are treated as major alarms although no loss of service results with the loss of only one converter.* The status signals indicate the spare modulators or demodulators in use. The command inputs are used to reset the transmission switches and to reset the SAC major and minor alarm outputs. All electronic alarm leads are brought individually into the SAC terminal strips to allow more detailed grouping if it becomes desirable later.

An important feature of the SAC is that it is ready to process another alarm immediately after it is reset, even though the first alarm is still in existence. "Reset" in this case means that the external indication of the first alarm is interrupted.

To set the SAC, a pulse, connected only during the initiation of an alarm, is used to set a bistable flip-flop. When the flip-flop is reset, it is immediately ready to recognize the next pulse. The resetting of an alarm does not cause a new alarm. Refer to Fig. 9; the pulse is formed by connecting the relay drive voltage through bunching contacts on the relay. When the relay is energized, this connects a negative pulse to the summary alarm circuit. This pulse breaks down the avalanche diode and sets the flip-flop. When the relay drive voltage is removed, the coil voltage immediately reverses and is clamped to +0.7 volt by CR1. The bunching of the contacts then connects a short portion of this voltage to the SAC. This voltage polarity does not set the flip-flop to the alarm condition. Since the lead to the SAC is connected only when the contacts bunch, there is no loading effect on pulses which might originate in other alarm circuits connected to the same flip-flop.

The flip-flop outputs are used to drive relays connected to the external alarm circuits. The transmitting and receiving alarm pulses are positive, but the operation is identical except that an NPN flip-flop is used. The dc-dc converters provide a permanent closure to the SAC. This closure is differentiated to generate a pulse which is processed in the same way.

The individual alarm lamps on the alarmed unit remain lighted when the SAC is reset so that the alarm information is not lost.

* The loss of one converter is considered to be a major alarm situation because of the serious hazard to service which exists in the bay in this condition.

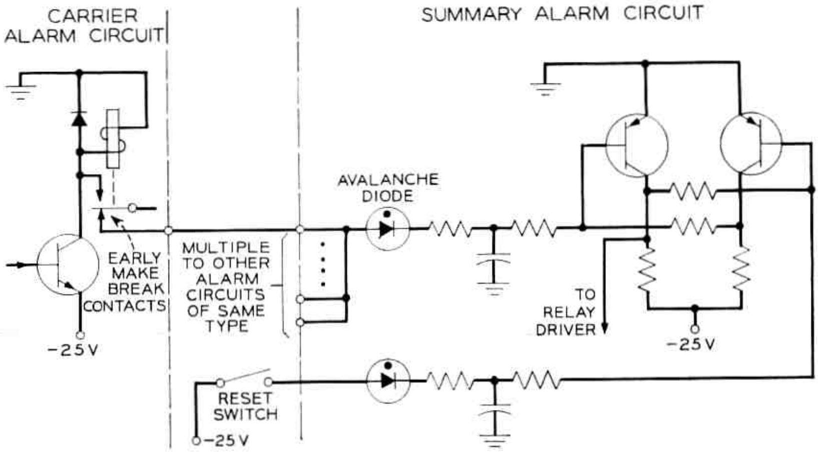

Fig. 9 — Typical alarm circuit.

The SAC is connected to two illuminated pushbuttons mounted in the front center of the bay. One is for major alarms, the other for minor alarms. The pushbuttons light when the SAC recognizes an alarm. When the button is pressed the SAC is reset.

2.4 *Physical Design*

The MMX-2 is a shop-assembled, -wired, and -tested unitized double bay which contains the transmission, carrier supply, switching, regulation, patching, and alarm equipment for a maximum of 18 mastergroups (10,800 channels). It also contains six spare mastergroup units that are automatically switched into service when needed. A maximum of three MMX-2 bays is required at an office terminating a 20-pipe (including standby) L-4 coaxial cable.

2.4.1 *General*

MMX-2 bays are generally located in a central office or a main station, in the maintenance aisle or test area where the aisle spacing is sufficient to permit the use of rolling test consoles without restricting the operating personnel. Because this area is the center of office activity, a good bay appearance, without significantly increasing the cost, was considered a design objective.

Seven-foot-high framework, which would be ideal for the long range trend to offices with lower ceilings, was considered. However, all projections indicate that, for the manufacturing life of the MMX-2, the

11 foot, 6 inch framework would make the most efficient use of space. Bays are available in 11 ft. 6 in., 10 ft. 6 in., and 9 ft. 0 in. arrangements.

To avoid excessive amounts of interbay cabling, the MMX-2 bay was designed as a complete mastergroup system incorporating the patching, test facilities, and equipment in a double unequal flange duct-type bay. By locating the transmitting and receiving equipment in the same bay assembly and by shop wiring, a consistent cabling and wiring plan was developed to avoid the interference and crosstalk paths often the result of a variety of field wiring and cabling techniques. In addition, the equipment has been compacted to the extent that it is no longer physically or economically practical to use field assembly and wiring methods.

The MMX-2 bay is shown in Fig. 10. A consistent modular design pattern for both shelves and modules has been developed throughout the bay. Meters indicating the phase of the carriers are mounted in the center, front uprights of the bay, adjacent to the carrier supplies being served. The space at the rear of the center duct contains the carrier transfer equipment. By using this space, which has not been used in the past, the carrier supply cable connections are very short.

All bays are completely shop-wired and assembled with shelves, even though the associated plug-in units may not be equipped. This does result in an initial higher cost; however, experience has shown that field additions would result in a much higher cost in the long run. It also minimizes the possibility of inadvertently interrupting working systems while the field additions are being made.

In addition to the front aisle, a wiring aisle is required at the rear of the bay. This provides two aisle surfaces for heat dissipation and better use of cabling space. The MMX-2 cannot be mounted in aisles back to back with other equipment; however, in view of some of the floor-loading restrictions, back-to-back equipment layouts in telephone offices have become less attractive as the equipment becomes more compact.

2.4.2 *Equipment Units*

The shelf assemblies were designed primarily as holders for the plug-in modules. Where practicable, the amount of equipment provided initially was restricted. Most modular units plug in but without fixed connectors on the shelves. In this way tolerance problems associated with mating connectors are avoided.

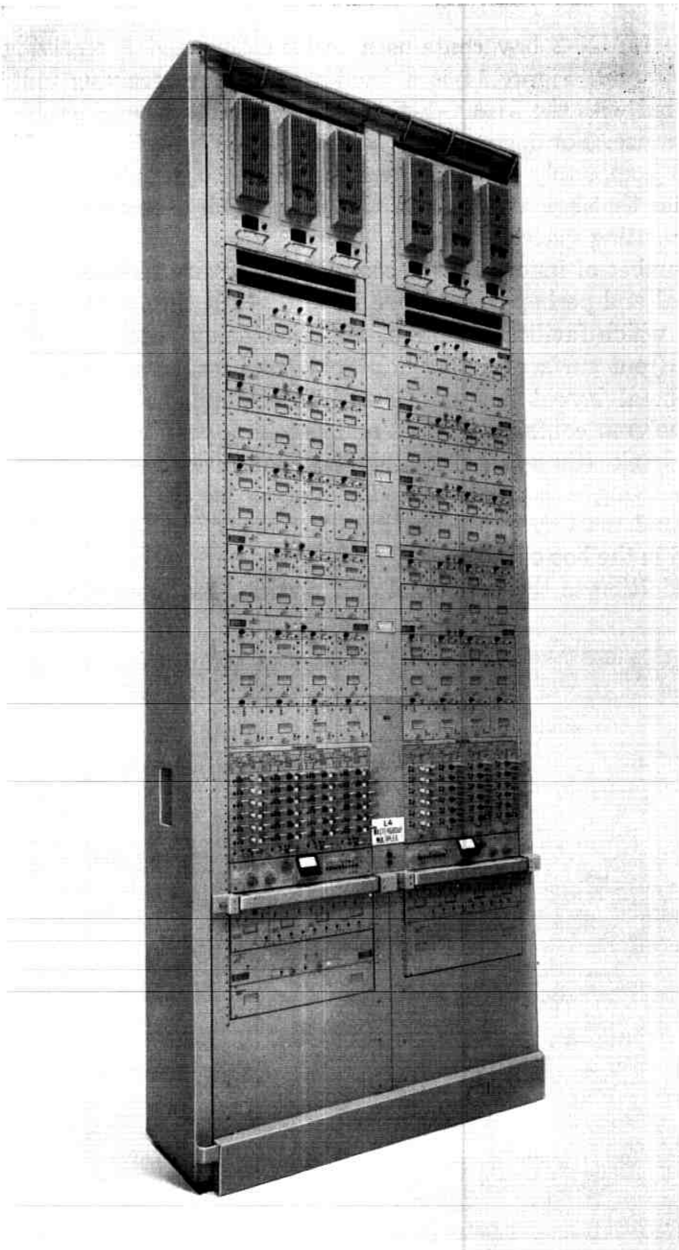

Fig. 10 — MMX-2 mastergroup multiplex.

2.4.3 *Patch and Test Unit*

Each MMX-2 bay contains a transmitting and a receiving patch and test unit. Figure 11 is a front view of the receiving unit. When compared with the MMX-1, this unit represents a considerable reduction in size. For instance, in MMX-1, a full 11 ft. 6 in. high bay almost completely equipped was required to provide the patching facilities that are available in this unit which uses only 12 inches of bay mounting space.

A number of innovations accomplished the size reduction. However, of equal and perhaps even more importance is the human engineering aspect which facilitates normal and emergency patching and testing.

The front surface of the jack field has been arranged to duplicate the general layout and position of the equipment units in the bay. Each column contains the test and patching position for one mastergroup bank (three regular and one spare), with each mastergroup bank serving a coaxial pipe. The horizontal lines separate the six individual mastergroups. The high frequency inputs and outputs are located in the bottom row.

In addition to the horizontal and vertical line separation, a general

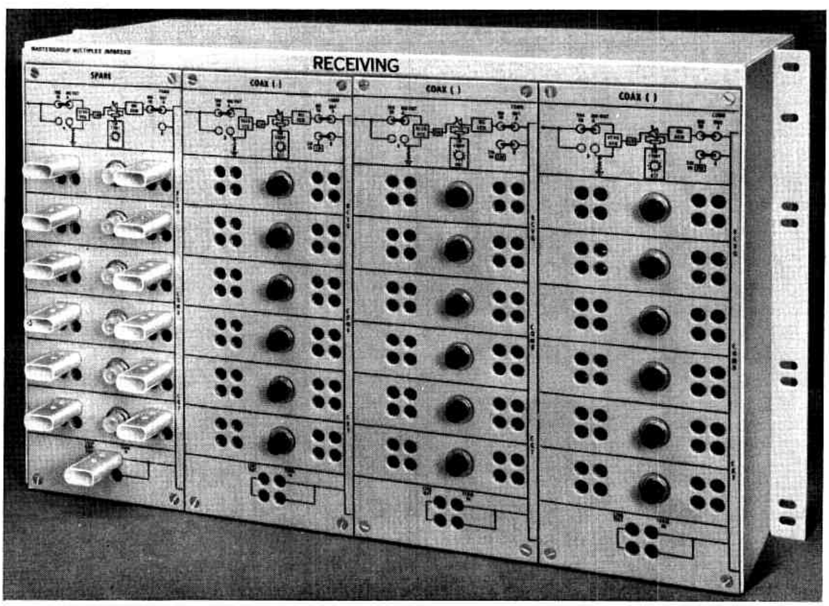

Fig. 11 — MMX-2 receiving patch and test unit.

block-type schematic is imprinted at the top of each column serving each mastergroup bank, giving the operating personnel a better understanding and means for associating the point in the system being patched.

The efficient use of the space available for the patching facilities was made possible by a complete new generation of Bell System coaxial jacks and plugs. They are scaled-down versions of the conventional reliable 477-type jacks and 356-type plugs. In addition to their use in the jack field, the new small coaxial jacks and plugs are used rather liberally for interconnecting apparatus within the bay.

Through the use of smaller transformers, it was possible to locate the input and protection hybrids and the combining and splitting networks to the rear of the jacks.

2.4.4 *Mastergroup Shelf Assembly*

Except for the mastergroup 1 shelf assembly, which does not have a carrier supply, the remaining shelf assemblies, for mastergroups 2 through 6, follow the same physical arrangement. Each MMX-2 bay contains six transmitting and six receiving shelf assemblies. Figure 12 shows a transmitting shelf assembly. A complete shelf assembly contains four (three regular and one spare) plug-in transmission modules and pilot detector modules.

The equipment for the carrier generator is located in the drawer which is hard-wired to the assembly. Apparatus can be replaced or maintained by extending the drawer on the built-in telescoping slides.

The front surface of the drawer is arranged to mount the lamps and switches required for transferring the carrier generator and also the lamps which indicate the status of the transmission modules. In order to accomplish the desired switch-lamp relationship in a reasonable space, new miniature lamps and switches were introduced. The new lamps, when illuminated, are sufficiently intense so that there is no difficulty locating them, even in a well-lighted environment.

The drawer is secured to the frame by a flush-mounted trigger-type latch which is used as a handle when released. This latch is rugged and provides a sure-lock that meets the MMX-2 shock and vibration requirements.

The framework for the shelf assembly is fabricated aluminum; the assembly is completely wired before being mounted in the bay. External connections are made, either at the terminal strips located on the sides of the shelf, or via coaxial cables spliced with 219-type connectors.

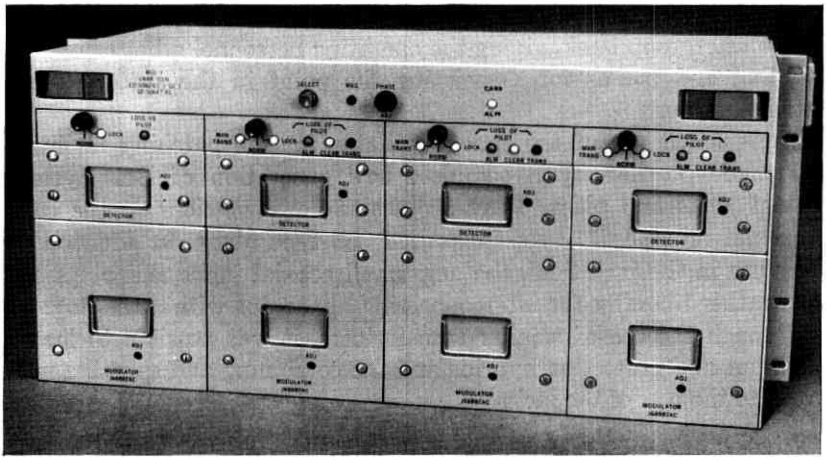

Fig. 12 — Mastergroup shelf assembly.

2.4.5 *Modulator Unit*

There are 24 modulator units in each fully equipped MMX-2 bay. Since mastergroup 1 does not require a step of modulation, the amplifier and detector apparatus are contained in a single plug-in type module. The modulator units for mastergroups 2 through 6 each contain two amplifiers, the modulator, a band-pass filter, and a low-pass filter. In order to achieve close-shielded connections, the above apparatus was connected by small coaxial plugs and jacks. Floating connections were avoided by sequential assembly and adequate clearance for the mounting holes to accommodate the normal tolerance variations.

For circuit reasons it was necessary that the input and output plugs of the band-pass filter be at opposite ends. In order to accommodate this requirement and still have a compact unit, the 266-type amplifiers that are connected to the band-pass filter are mounted in an upright position. This has the added benefit that the heat-producing amplifiers are on opposite surfaces of the MMX-2 bay.

The demodulator unit, which reverses the transmission process, uses essentially the same apparatus packaged in a similar type module.

2.4.6 *Carrier Supply Amplifier*

To avoid congestion or the possibility of a larger drawer, a monolithic crystal filter was developed and mounted as a part of the carrier amplifier shown in Fig. 13. The amplifier is secured to the base of the drawer with the input and output connections made through coaxial jacks and plugs. The amplifier components and the monolithic crystal filter are mounted on a printed wiring board. In order to maintain the high input-to-output loss, a special copper shield was required between the terminals of the filters.

The physical design advantages of MMX-2 over MMX-1 are:

(*i*) Floor space savings of better than 8:1 are realized.

(*ii*) Modules are equipped as required for service.

(*iii*) Service interruptions resulting from failure are significantly reduced by the automatic transmission and carrier switching.

(*iv*) Field testing and maintenance have been facilitated.

(*v*) Intraoffice cabling is reduced by intrabay shop cabling.

(*vi*) Complete flexibility for route assignment has been achieved.

(*vii*) The installation-to-service interval is reduced.

2.5 *Filters and Networks*

Frequency multiplexing, by its very nature, is a large user of filters and networks, such as band-pass filters, carrier and pilot selection filters, and band elimination filters. A variety of techniques and components are used.

Much of the L-4 networks' design and development challenge came because of three factors: the bandwidths are large, the transition bands narrow (4 percent) but adequate, and a large portion of the L-4 frequency spectrum falls in the awkward region for ferrite and air core inductors; that is, above 3 MHz and below 25 MHz. Ferrite

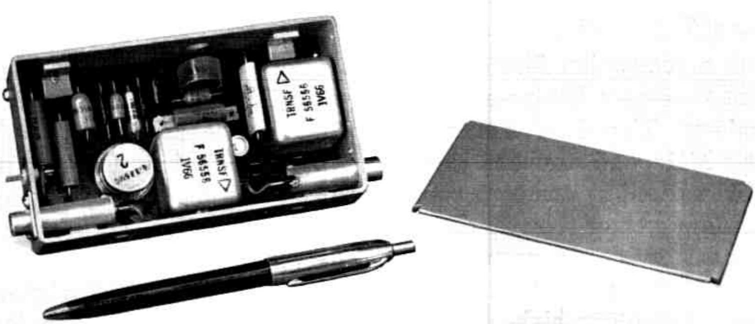

Fig. 13 — Carrier amplifier.

components are used wherever possible because of their generally superior Q, self shielding, and smaller size. Because of the extensive use of pilots and single-frequency tones in the L-4 system, about 50 percent of all of the filter designs required are crystal filters.

2.5.1 *Transmitting and Receiving Band-Pass Filters*

The L-4 mastergroup multiplex terminal requires a set of wide-band sharp cutoff band-pass filters: a transmitting (except for MG-1) and a receiving filter for each mastergroup. After analyzing the stop-band loss objectives for each mastergroup's transmitting and receiving filter (see Section 2.1.3), it was determined that the same design could be used for both. The mastergroup 1 receiving band-pass filter, with a bandwidth greater than 100 percent, is configured as a highpass-lowpass elliptic function filter of degree 9 and 12, respectively, with a 3 percent reflection coefficient for good inband impedance. High Q ferrite inductors are used in this band-pass filter. Mastergroup 2 through 6 band-pass filters use air core inductors, each mounted in a separate can with the rest of the filter section components. During the design and development stage, care was taken to correct for the large number of parasitics encountered. An insertion loss design of degree 22 proved satisfactory for mastergroup 2. Mastergroup band-pass filters 3 through 6 are image designs to facilitate development and tuning, with a newly developed terminating end section for parasitic correction, good image impedance and element values. Because of the effects of Q, each filter has an integral bridged-T amplitude equalizer. Figure 14 is a typical schematic diagram with the loss and return loss characteristics of a mastergroup band-pass filter.

2.5.2 *Transmitting and Receiving Low-Pass Filter*

As a companion filter to the mastergroup band-pass filter, a constant resistance low-pass filter is used to limit noise and modulation products. This filter has low distortion in the passband, 0.564 to 3.084 MHz, and stopband loss greater than 65 dB from 6.336 to 100.0 MHz, as well as constant resistance input impedance properties (return loss greater than 26 dB) from 0.564 to 35 MHz. To achieve the constant resistance properties in the stop band the low-pass filter was designed so that an impedance correcting network consisting of a complementary high-pass filter terminated in 75 ohms, the design impedance, could be used. The high-pass network contains fewer sections, 2 versus 4 for the low-pass network, because the system ob-

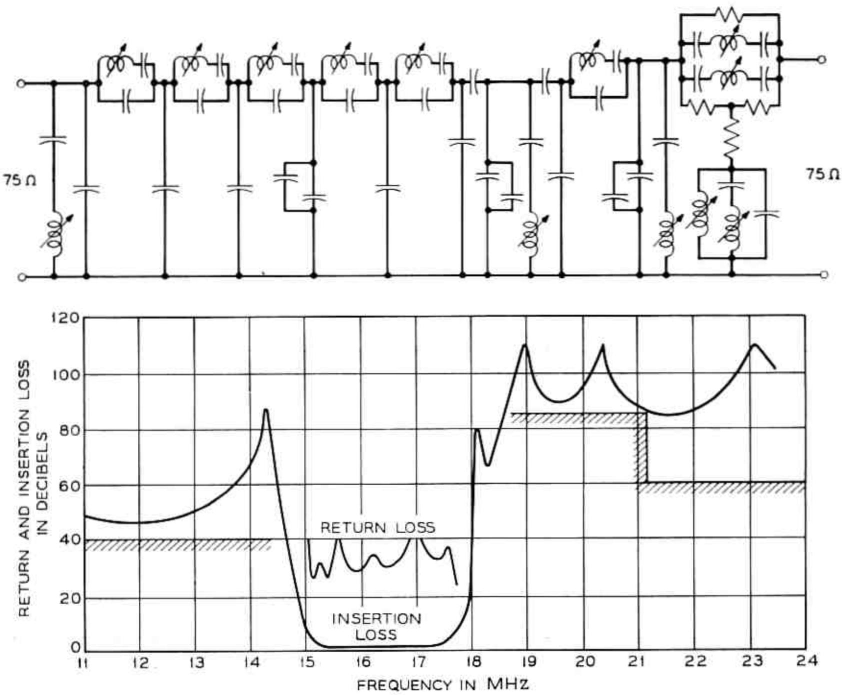

Fig. 14 — Schematic and loss versus frequency characteristic of typical MMX-2 transmitting and receiving band-pass filter.

jectives could be achieved by using an approximately complementary network, thus realizing some component economies. Small air core inductors with individual shields are used because of their stability and large useful frequency range. The disadvantage of low Q is not a great penalty; the small slope distortion across the band is corrected by a simple bridged-T equalizer.

2.5.3 *Carrier Supply Band-Pass Filters*

The L-4 carrier supply requires 21 narrow band-pass filter designs. Eight are used in the binary frequency supply. These are simple LC filters with midfrequencies (f_m) from 128 kHz to 16,384 kHz in binary multiples, 10 percent bandwidth, and more than 40 dB discrimination to $f_m/2$ frequencies. Another eight filter designs select the proper frequencies in the carrier generator. These are also LC filters, for the most part of greater complexity than the binary supply filters. All 16 filters are constructed with small shielded air core inductors mounted on printed wiring boards sealed in individual cans.

One of the more interesting and most unusual filter designs is the monolithic crystal filter (see Fig. 15) used in the carrier supply amplifier.[4, 5] An extremely narrow band with stopband losses in excess of 80 dB is achieved in a small volume, less than one-half cubic inch, crystal-only filter without the use of any additional elements.

III. LINE CONNECTING CIRCUITS

As discussed in the introduction, the line connecting circuits are the means by which the L-4 line frequency signal is interconnected and controlled within a main station. Figure 16 shows the line connecting circuits in block diagram form. At a terminal point, these circuits interconnect the MMX-2 with the L-4 line. At through message stations, these circuits allow connection of all, or a selected part, of the received line signal to the transmitting line. In addition, the line connecting circuit provides line pilots, and controls the transmission or blocking of the various test and control tones required by the L-4 system through the main station.

One of the major influences in the design of the line connecting circuits was the initial decision that automatic protection would be provided for all active circuitry. While the transistor circuitry alone might have been considered sufficiently reliable not to require protection, the necessary power wiring, fusing, and power supplies, in addition to the basic circuitry with its susceptibility to operating error, constituted an unacceptable service hazard. As a consequence of this basic decision, it was determined that the major transmission path through the office should, if possible, be made completely passive. This requires a minimum of equipment in a through connection and results in better performance and higher reliability.

3.1 *Receiving and Transmitting Hybrid Networks*

The receiving and transmitting portions of the line connecting circuits contain hybrid networks which perform required signal splitting and combining functions. The receiving hybrid network splits the incoming line signal and provides five outputs at different levels. The lowest loss path through the network provides a connection to the blocking circuit, the major through connection in the office. The other outputs may be connected to: the receiving MMX-2 connecting circuit for dropping, the branching circuit, the sync receiving circuit, and a

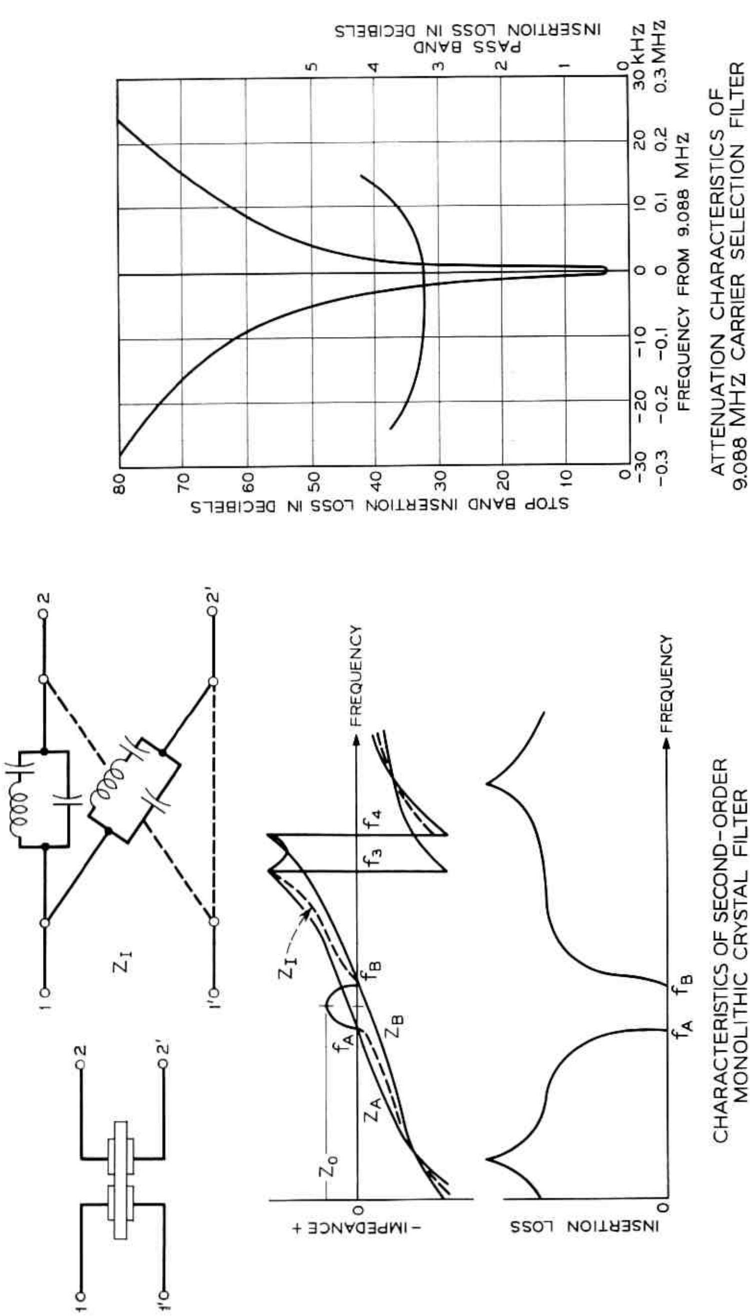

Fig. 15 — Monolithic crystal filter.

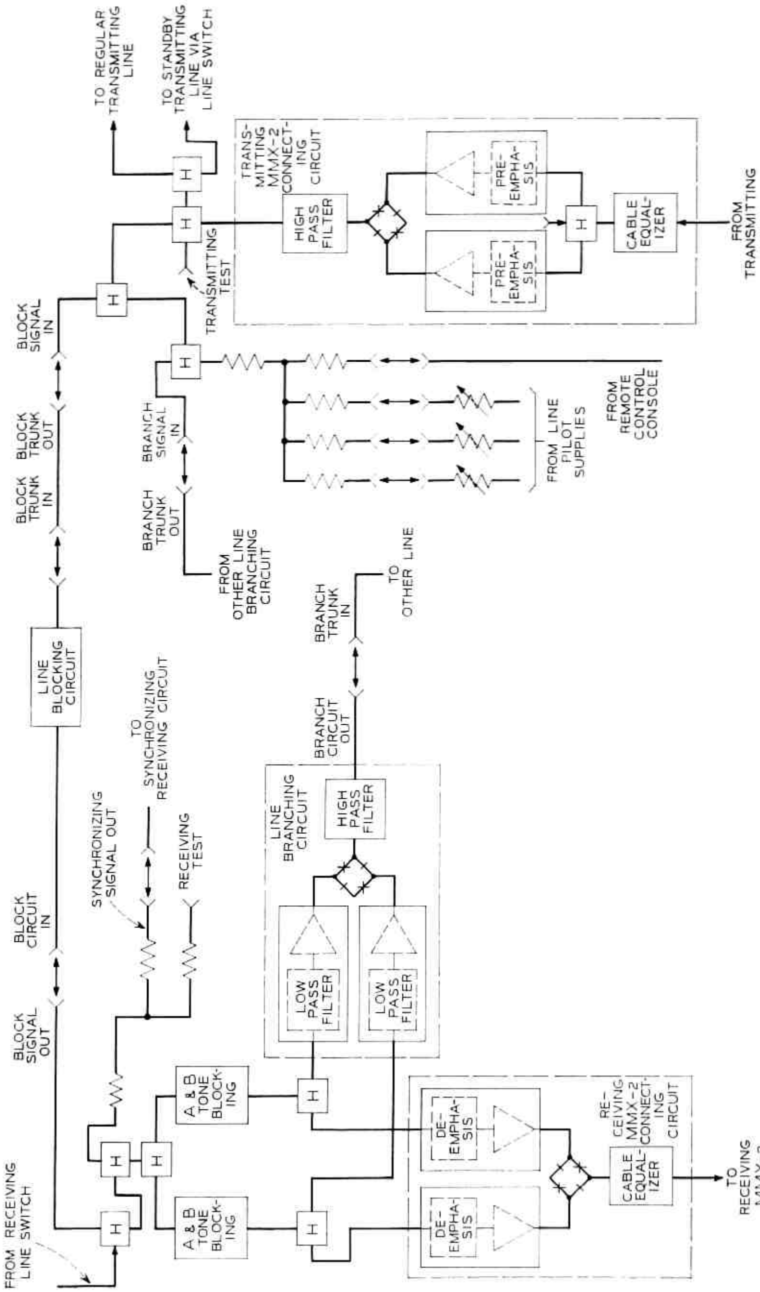

Fig. 16 — Line connecting circuits.

receiving test jack. The transmitting hybrid circuit accepts inputs from the blocking circuit, the branching circuit, the transmitting MMX-2 connecting circuit, the line pilot supplies, and the remote control center. Three outputs of the combined signal are derived. One output is connected to the transmitting line amplifier for the regular line. A second is connected into the transmitting line switching circuit where it can be connected into the protection line when required. The third output provides test access to the combined signal.

3.2 *Line Pilot Supplies*

There are three full-time pilots: 512 kHz for synchronization and low-edge regulation, 11.648 MHz for a line regulating pilot, and 20.448 MHz for the high-edge pilot. The two higher frequencies are generated in free-running crystal oscillators. The 512 kHz is obtained from the office primary frequency supply. Each office primary frequency supply can be synchronized by the incoming 512 kHz pilot.[6]

The two higher frequency supplies are almost identical, except for operating frequency, and are very similar to the 2.840 MHz mastergroup pilot supply in the MMX-2 bay. The major difference between the mastergroup and line pilot supplies is that each of the ten outputs from the distribution circuit for the line pilots can be connected through relays for local pilot insertion.

The 512 kHz supply's balanced input at −23 dBm and 135 ohms is processed by a clipping stabilizer, filtered and distributed. The clipping stabilizer has a 10 dB input dynamic range with only a 0.1 dB output change.

The clipper is a temperature-compensated diode circuit similar to the one used in the pilot oscillators. An input detector is used to change the bias in the clipper amplifier to cut it off if the input drops by more than 10 dB. This prevents placing a potentially noisy synchronizing pilot on the line.

The 512 kHz supply is redundant in the same way as all the other pilot supplies. The stabilizer alarm detectors, however, are "window" detectors which alarm when the output signals deviate in either direction by more than 0.5 dB from nominal. This is necessary since the stabilizers are very stiff, and only slight output errors must be detected in order to properly recognize an out-of-service condition.

3.3 *Synchronization Receive Circuit*

The synchronizing pilot at 512 kHz is obtained from the same splitting pad which provides the signal to the receive test jack. The power at this point is approximately −52 dBm, which must be amplified before the signal is distributed to the office primary frequency supplies. This function is accomplished in the 512 kHz synchronization receiving circuit. This panel provides three output taps at −23 dBm and 135 ohms balanced. The number of taps was intentionally limited to prevent connection of more than three primary frequency supplies.*

The input impedance of the synchronization receiving circuit must have a return loss in excess of 26 dB to prevent disturbing measurements made at the receive test jack. Also, it must accept the entire line spectrum. This necessitates a pad and amplifier at the input prior to the crystal 512 kHz pickoff filter. Thus the higher gain amplifier following the filter does not require a high level output capability, nor is the filter subjected to a high input level.

This panel is not redundant and has no alarm outputs. The line alarms adequately indicate the loss of the 512 kHz pilot and redundancy is not necessary since the primary frequency supplies can free run for a short time with little system impairment. Instead of redundancy, a mounting for a cold spare unit is provided in the panel. This prealigned cold spare can be manually placed in service in a short time.

3.4 *Line Blocking Circuit*

The blocking circuit's primary function is that of connecting only desired portions of the L-4 line signal from the receiving circuits to the transmitting circuits. The signal which is transmitted may consist of all or part of the message spectrum, line pilots and, under some circumstances, a part of the equalizer control signals.

In the early stages of developing the blocking circuits it was recognized that, although the ultimate in message signal blocking would be to have completely unrestricted blocking, there would be advantages if something less than this were acceptable. Consultation with the operating companies revealed that the more restricted plan of requiring only contiguous sets of mastergroups to be blocked was adequate.

By restricting the message blocking capability to only sets of contiguous mastergroups, a much more manageable system was made possible. Instead of using tandem connected band elimination filters, it is possible to use a parallel combination of a high-pass and a low-

* It is intended that there be one office master (and an alternate) primary frequency supply from which all other primary frequency supplies are synchronized.

pass filter which have overlapping stopbands to block the message signal. Consequently, the passed signal encounters the distortion and loss introduced by only one filter. Section 3.9.1 gives an example of blocking. The flat loss is 10 dB: 8 dB for the blocking filters, and 1 dB each for the splitting and combining filters. The total flat loss in the passband of the complete blocking circuit is 15.9 dB.

The L-4 system design allows the line pilots to be transmitted through as much as a complete frogging section, 800 miles maximum, without being blocked and reinserted. However, it is not a requirement that this be done. It is permissible to block and reinsert more often if necessary.

The 512 kHz line pilot is treated differently from the 11.648 and 20.448 MHz pilots because of its use as a synchronizing signal as well as a line pilot. In order to ensure good synchronizing, it was originally decided to transmit the 512 kHz pilot through the full system without blocking.* Each station bridges the synchronizing signal off the line in L-4. This is in contrast to earlier systems which blocked and reinserted frequently, sometimes at every main station, thus introducing onto the pilot the noise generated by the office synchronizing equipment.

In order to allow transmission of the 512 kHz signal over a full system, the line blocking circuit was designed to pass 512 kHz in all of its applications. For example, when mastergroup 1 is blocked, a 512 kHz band-pass filter is paralleled with the blocking filter. In addition, a blocking circuit which passes only 512 kHz was designed for use at points where the full message load is dropped.

The 11.648 MHz line pilot is blocked and reinserted whenever mastergroups 4 or 5 are blocked. When both are blocked, the message blocking filters remove the pilot as well. When one but not the other is blocked, some of the pilot signal will leak through, since the blocking filter is in its transition region at this frequency. In this case, a pilot blocking filter is added which suppresses the pilot by a minimum of 50 dB after which the pilot is reinserted. The 20.448 MHz pilot is blocked when mastergroup 6 is blocked. No additional pilot blocking filter is required since the message blocking will always suppress the pilot to the required extent.

Two other sets of signals are controlled by the line blocking circuit: the line switching tones and the equalizer control tones. The

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* For reasons beyond the scope of this paper, the 512 kHz pilot is now blocked and reinserted at frogging points (approximately 800-mile intervals).

switching tones must be suppressed at every main station to avoid inter-section switching interactions. All of the equalizer control tones must be suppressed at main stations that have remote control centers. At stations without remote control centers, only the lower half must be blocked in one direction of transmission and only the upper half in the other direction.

Three filters have been designed to meet these requirements. At stations which require complete blocking, a 512 kHz highpass filter is used in the blocking circuit which suppresses the switching and control tones in the 280 to 500 kHz band but passes the 512 kHz pilot. When partial blocking is required, one direction of transmission uses a 400 kHz high-pass filter which blocks only the lower half of the control tones along with the switching tones. In the other direction, a complex filter is used which suppresses the switching tones and the upper half of the equalizer tones between 400 and 500 kHz.

Another version of the line blocking circuit is used with the protection line. This circuit is intended for use at main stations which do not have remote control centers and allows the equalizing test tones to pass through the office to the next main station, which must have a remote control center. Without this circuit, it would not be possible to equalize the far main section of the protection line when two-section equalizer control is required.[3] In addition, all line pilots, line switching tones, and equalizer control tones are blocked.

3.5 A *and* B *Equalizer Tone Blocking Circuit*

At the inception of the L-4 system design, it was planned to provide A and B equalizer test tone blocking in the main transmission path at every main station which was equipped with a remote-control center. This would isolate the equalizing sections from each other and obviate any need for coordination between remote-control consoles. As the design of the system and of the required blocking filters progressed, it became apparent that this was not practicable. The complexity of the filter design was such that ultimately eight filters were required, with a combined loss to the passed signal of approximately 59 dB exclusive of padding needed to avoid impedance interactions. Inclusion of this large number of networks with the necessary amplification in each main station with a remote control center would have seriously strained the L-4 system amplitude and phase distortion requirements. Therefore, it was decided to place the tone blocking circuit in the dropping and branching paths. Consequently, a through mastergroup encounters only one of these circuits in a frogging section, at the frogging point where it is dropped. In addition, it prevents interaction between systems by way of branching or interconnection through mastergroup connectors. Since the tones are passed through the blocking circuits within a frogging section, coordination between remote control centers is required in the frogging section.

Two complete tone blocking circuits are provided. Each consists of tandem connected tone blocking filters, amplifiers, and impedance isolating pads. The two circuits are fed from identical outputs from the receiving hybrid circuit. Each side has a hybrid circuit as its final component which provides an output to the receiving MMX-2 connecting circuit and another to the branching circuit. One side or the other is selected in the connecting and branching circuits.

3.6 *Receiving MMX-2 Connecting Circuit*

The receiving MMX-2 connecting circuit is composed of parallel transmission paths, each containing a de-emphasis network, an amplifier, and a hybrid. The hybrid on each side feeds an output to a coaxial switch and an output to a pilot detector circuit. The pilot detector circuit recognizes the presence or absence of the 512 kHz line pilot and operates the coaxial switch to select one of the transmission paths from the receiving hybrid circuit, through the A and B tone blocking circuits, and through the connecting circuit. The switching logic and power fusing have been arranged such that loss of a single main discharge fuse will not cause a loss of service.The logic is also arranged such that complete loss of the incoming pilot from the line will not cause a switch. The output of the coaxial switch is fed through a cable equalizer to the receiving MMX-2 terminal. The equalizer is capable of equalizing for up to 200 feet of cable between the receiving line bay and the MMX-2 bay.

3.7 *Line Branching Circuit*

The branching circuit is the complement of the blocking circuit. While the blocking circuit blocks sets of contiguous mastergroups, the branching circuit passes sets of contiguous mastergroups. The blocking circuit function is performed by parallel high-pass and low-pass filters with overlapping rejection bands. The branching function is performed by these same filters but having overlapping passbands and operated in tandem. While the blocking circuit often passes line pilots, the branching circuit never passes them.

The branching circuit requires the use of an amplifier because of the high loss introduced by the tandem connection of the message blocking high-pass and low-pass filters, 16 dB, as well as the pilot blocking filters and cable equalizer. Because of this, it contains a transmission switching circuit nearly identical to that used in the receiving MMX-2 connecting circuit.

3.8 *Transmitting MMX-2 Connecting Circuit*

The transmitting MMX-2 connecting circuit is a dual switched circuit which prepares the output signal of the MMX-2 bay for interconnection into the L-4 line by applying the required pre-emphasis and adjusting its level. The switching circuit is identical to that used in the receiving MMX-2 connecting circuit and the branching circuit. A 512 kHz signal is applied at the input of the circuit and used to operate the switching. A 544 kHz high-pass filter is used at the switch output to remove the 512 kHz signal as well as any undesirable low frequency noise originating in the MMX-2.

3.9 *Line Connecting Filters and Networks*

From a network point of view, the line connecting function is dominated by two very difficult filter designs: the mastergroup blocking and branching filters and the A and B tone blocking filters.

3.9.1 *Mastergroup Blocking and Branching Filters*

The blocking of contiguous mastergroups is achieved by a combination of one high-pass, one low-pass, and two split-apart filters. Five designs of each high-pass, low-pass, and split-apart filter are required to accomplish this versatility. For example, to block mastergroups 4 and 5, a low-pass filter that passes mastergroups 1, 2 and 3, a high-pass filter that passes mastergroup 6, and a pair of split-apart filters with transition regions in mastergroup 4 or 5 are required. The use of the more complex split-apart filters instead of a pair of hybrids gives two important advantages: the flat loss is reduced to 2 dB (1 dB per split-apart filter) versus 6 dB for a pair of hybrids, and more important, the inband return loss is good. The inband return loss would not be good in the case of hybrid use.

The low-pass and high-pass filters are insertion loss designs, each of degree 19. The split-apart filters are also insertion loss designs but of degree 6. A considerable amount of loss distortion, up to 6 dB, at the passband edge is a result of the steep transition band and the ferrite inductor Q of approximately 200. This distortion is equalized by bridged-T equalizers which are an integral part of each low-pass and high-pass filter. Figure 17 is a typical schematic diagram for a mastergroup blocking filter arrangement.

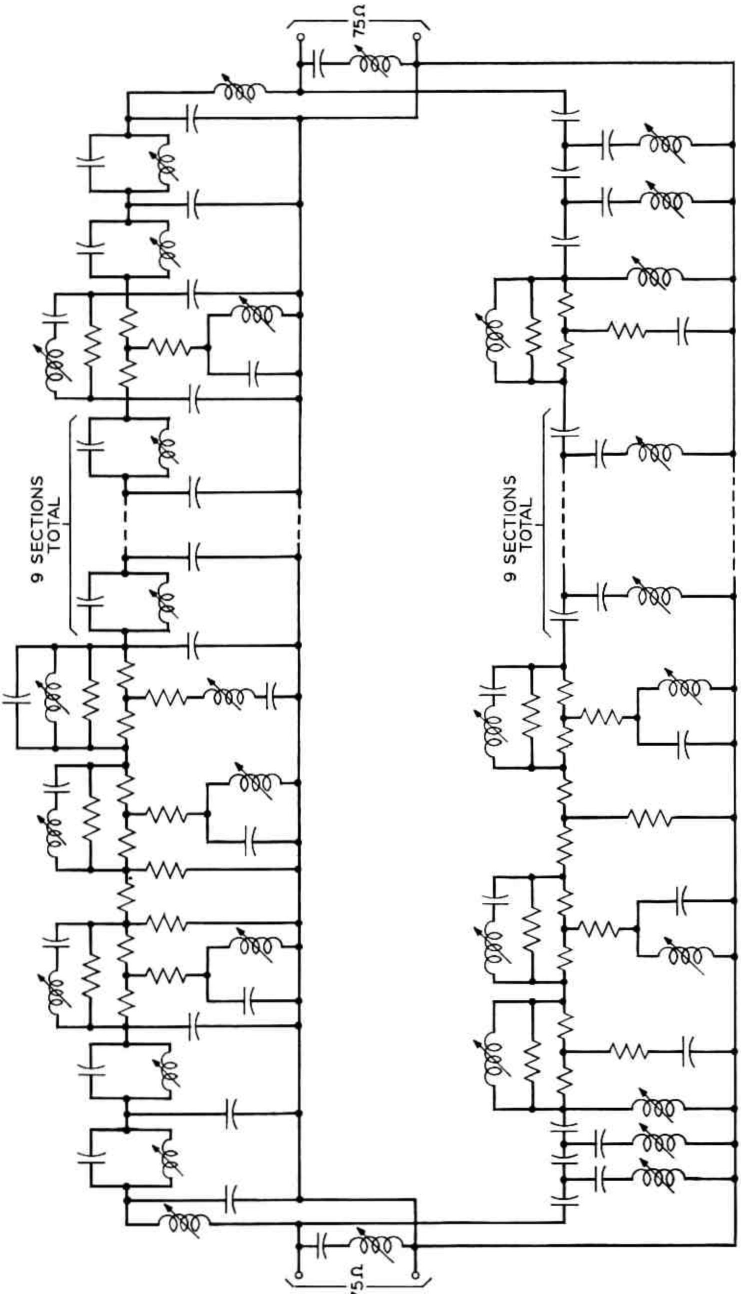

Fig. 17 — Schematic of typical mastergroup blocking filter arrangement.

Considerable care was taken during development of these filters to compensate for or reduce the parasitic capacitances, as well as to eliminate ground loops and other coupling paths. These filters must maintain discriminations in excess of 80 dB in the blocked mastergroups.

These same high-pass and low-pass filters (without the split-apart filters) are used in the mastergroup branching circuit. The appropriate high-pass and low-pass filters are chosen and put in tandem to form a band-pass filter to branch the mastergroups desired.

3.9.2 *512 kHz, 11.648 MHz, and 20.448 MHz Pilot Blocking Filters*

The blocking and branching circuits also contain the blocking filters for the three full-time line pilots, 512 kHz and 11.648 and 20.448 MHz. The 512 kHz pilot is blocked by a high-pass filter with a passband from 544 kHz to 21 MHz, suppressing 512 kHz by a minimum of 70 dB. The 544 kHz high-pass filter was designed with an equal ripple passband and arbitrary stopband with six finite-frequency poles and a pole at zero. Computer programs were used for the pole placement as well as for the synthesis. The severe objective of passband flatness, 0.1 dB up to 21 MHz, required extreme attention to detail during the physical realization of the filter.

The 11.648 and 20.448 MHz pilot blocking filters are conventional Dagnall crystal band-elimination filters. Both use the same electrical and mechanical structure; only the crystals are different. Since these band-elimination filters are always in the main transmission path, care was exercised to reduce all unwanted crystal responses. By using the trapped energy principle during crystal design, and AT cut crystals, the unwanted responses were reduced to an acceptably low level.

3.9.3 *Command, Line Switching, and Monitoring Tone Blocking Filters*

Command tones in the 300 to 500 kHz range, and monitoring tones in the 18.500 to 18.560 MHz range, must be blocked and passed in various combinations as outlined in Section 3.4. A scaled version of the 544 kHz high-pass filter reviewed in Section 3.9.2, designated the 512 kHz high-pass filter, passes from 512 kHz to 21 MHz and suppresses the command signals, 300 to 500 kHz, by a minimum of 40 dB. This filter is in turn combined within a complex network made up of input and output split-apart terminations with a band-pass filter, 316 kHz–396 kHz, an equalizer in the upper leg and the 512 kHz high-pass filter and equalizer in the lower leg (see Fig. 18). The overall characteristic of this command channel subgroup II filter passes the basic L-4 transmission spectrum with the 512 kHz synchronizing

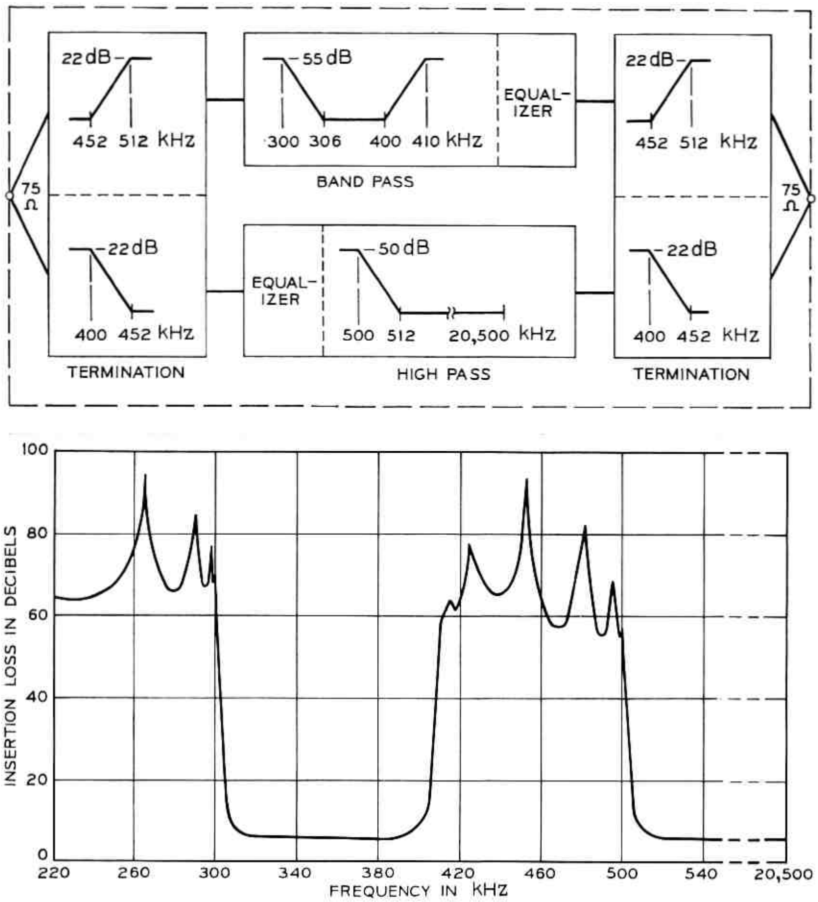

Fig. 18 — Command and line switching tone blocking filter.

pilot and passes subgroup II command channel tones while rejecting the line switching tones below 300 kHz and the rest of the command channel tones.

The sixteen monitoring oscillator tones, 18.500 to 18.560 MHz, are blocked by an LC band-elimination filter using a pair of special band-elimination sections and incorporating infinite-Q compensation.[7]

3.9.4 A *and* B *Equalizer Tone Blocking Filters*

The problem of suppressing a large number of single frequency tones, used for control of the A and B equalizers, by a minimum of 50 dB and passing all the L-4 message bands with less than 0.1 dB distortion had to be solved. Three design approaches were used. For frequencies 5.888 MHz to the highest B tone, 20.200 MHz (where the third overtone of a crystal resonating at the tone frequency falls above the passband of interest), a Dagnall type filter, combining a low-pass filter with shunting quartz crystals, is used. For frequencies 1.056 to 4.256 MHz (where the third overtone falls in the L-4 transmission band), the more complex band-pass-band-elimination filter principle is used. At the lowest tone frequency, 0.544 MHz, where the transition bandwidths are more generous, a conventional LC band-elimination filter using high Q ferrite inductors is used.

The Dagnall high frequency tone blocking filters consist of a number of low-pass filter sections, whose cutoffs are above the highest frequency in the L-4 band, with the shunt capacitance replaced by a crystal(s) and fixed capacitor. The crystals at series resonance shunt energy to ground, thus giving rise to the band-elimination effect.

At the low A and B tone frequencies, where the third overtones fall in the L-4 band and the unwanted responses were a problem, a considerably more complex design was used. Figure 19 outlines the complexity of a typical low frequency A or B tone band-elimination filter. The input and output band-pass and band-elimination split-apart filters were designed as complementary pairs with approximately 20 to 40 percent bandwidths; then a crystal band-elimination filter is inserted in the band-pass leg. The design of Fig. 19 consists of two

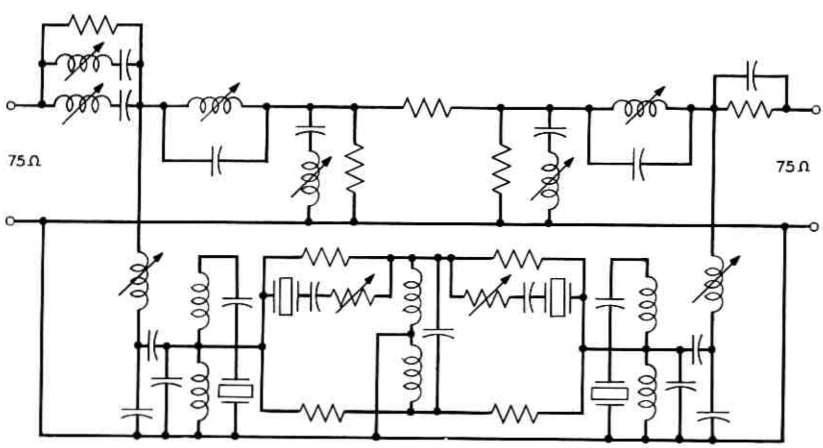

Fig. 19 — Schematic of typical low frequency A- or B-tone band-elimination filter.

crystals shunting the high impedance autotransformers and two special lattice bridge sections. These half-lattice bridge sections are low loss pads with a crystal shunting the large resistor. At crystal resonance, the effective resistance of the crystal shunts the high resistance leg, forming a bridge balance and a transmission null. This design is used to achieve sufficient suppression magnitude and bandwidth without the penalty of large insertion phase at passband frequencies. The network in the band-elimination leg compensates for the insertion loss and phase in the band-pass leg. The flat loss of one of these filters is nominally 8 to 9 dB.

IV. MASTERGROUP CONNECTORS

4.1 *Early Connectors*

Although in principle every message channel could be demultiplexed all the way to voice frequency level at a coaxial system main station, this is seldom done in practice. Many 12-channel groups and 60-channel supergroups are conveyed across the office without further demultiplexing, and fed to other coaxial or radio routes.

The filtering and transmission level adjusting circuits between systems are known as group or supergroup connectors. They provide economies and performance improvement by virtue of reducing the number of channel and group banks in a long distance telephone circuit. They also make possible the routing of group and supergroup bandwidth data circuits from system to system without spoiling the integrity of the wide frequency band.

In 1963 a mastergroup connector became available, capable of interconnecting basic mastergroups among L-3 coaxial and TH (and some TD-2) radio systems, specifically by interconnecting the MMX-1 terminals.

4.2 *Mastergroup Connectors for L-4*

New mastergroup connectors were designed for three reasons. New connectors were desirable to exploit the better band-pass filter in-band frequency characteristic made possible by the more generous spacing of MMX-2 derived mastergroups. New connectors were necessary to suppress the 2.840 MHz pilot in the MMX-2 derived mastergroup to prevent its interference with pilots on immediately connected or remote systems. Finally, field experience with the MMX-1 to MMX-1 connector indicated that a single equipment package containing the circuits for both directions of transmission would be desirable.

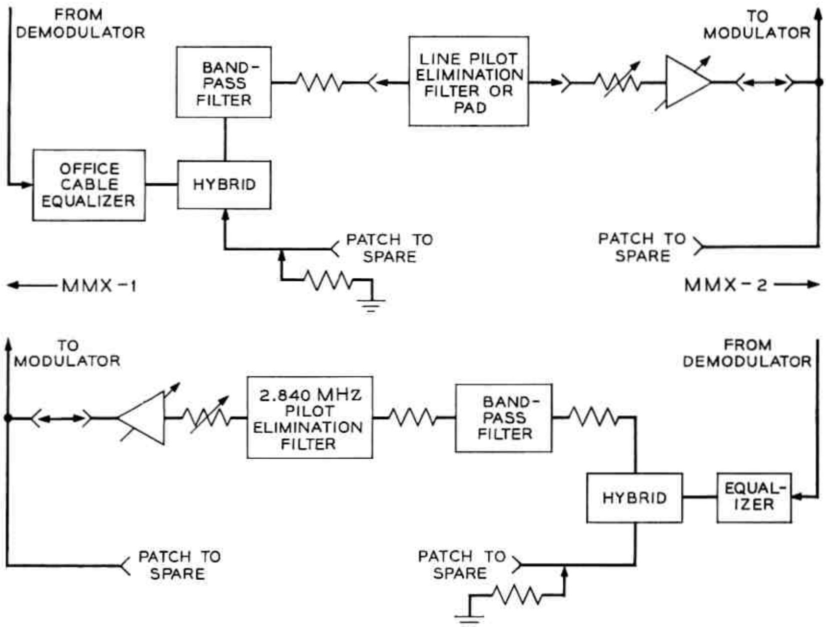

Fig. 20 — Mastergroup connector, MMX-1 to MMX-2 (two-way).

Consequently, two new two-way mastergroup connectors were developed. Figure 20 shows the MMX-1 to MMX-2 connector. The MMX-2 to MMX-2 connector is similar; both halves look like the bottom part of Fig. 20.

V. LINE PROTECTION SWITCHING

5.1 *L-4 Protection Switching*

One pair of coaxials on an L-4 route is assigned as a protection pair for the lines in service. The number of pairs in service is therefore nine for a fully equipped 20-coaxial cable route. To help meet early L-4 service schedules, the already available L-3 protection switching system was modified for use on L-4 routes.[8] The modification has two essential aspects:

(*i*) The L-4 pilots, used to signal the condition of the line to the switching control system, differ in number and frequency from those in L-3.

(*ii*) The modified system has been expanded to operate on a one-for-nine basis. Need for this modification is really a result of the use of the 20-coaxial cable instead of the older 12-coaxial or 8-coaxial cable. In principle, the expansion could also be used with L-3.

5.2 *Switching Control System*

The L-3/L-4 switching control system, developed in the early 1950's for L-3, depends on transmission of two types of signal. One consists of the line pilots (512 kHz and 11.648 MHz for L-4) which cause switch initiation if their power levels change substantially. The other is a revertive signal which goes back over opposite-direction lines to the transmitting office to control the transmitting switch. This signal consists of tones in the 280 to 290 kHz band. Finally, a 296 kHz forward-transmitted tone verifies operation of the transmitting switch and thus signals the receiving switch to operate.

The receiving control circuits are provided on each line and have access to the particular line via a hybrid in the receiving line connecting circuits, as suggested in Fig. 2. Each of the two line pilots is selected by a pick-off filter, amplified and rectified in a detector circuit. Each detector circuit drives a control microammeter. If any of the two control meters deviates more than 2 dB from zero, it activates office alarms, lights the alarm lamp associated with the meter, and initiates a "slow" switch, so called because the meter can take as much as a quarter of a second to swing past the 2 dB limit and initiate a switch.

The slow switch is intended for automatically switching out a badly misequalized line, or one for which a control pilot has drifted beyond 2 dB for any reason. In case of a catastrophic failure the 11.648 MHz detector provides, besides the meter drive signal, a dc signal which bypasses the meter and initiates switching action as soon as the pilot deviates plus 4 or minus 5 dB. The resulting "fast switch" is completed in about 15 milliseconds after the failure.

5.3 *Coaxial Switches*

Three types of coaxial switch are used in L-4: a 223B for the transmitting switch, 223C for the line terminating switch* and 223D for the receiving switch. To minimize the transient caused by a manual line switch, the 223D switch was designed as a make-before-break switch, using the 237 type sealed reed element for the contacts. On a manual switch there is no change in the regular line path until the receiving

* In case of total failure, the line terminating switch terminates the incoming line.

switch operates; the transmitting switch, which operates first, merely puts the message signal on the standby line as well as on the regular line. The "make" contacts are closed by an electronic timing circuit, which is an integral part of the 223D switch, before the "break" contacts are opened by the timing circuit. On switching back, the "break" contacts are closed first. In a multiline switching system, one to four regular lines are equipped (per direction). The switching control bays are connected so that these tones go back over two to five lines, including the standby line.

An "expanded multiline" system, good for up to eight lines (per direction), can be installed by connecting the four tones to the return lines in two groups; this adds space-diversity to the frequency diversity so that one out of eight lines can be identified by the tone frequency and the group it is found on at the transmitting station.

For 20-coaxial use, particularly for the early L-4 installations, further modifications provide "super-expanded multi-line" capability by extending the line grouping to three groups. This allows up to one-for-twelve switching, but one-for-nine switching is expected to be the maximum used in the immediate future.

Such expansions of multiline capacity by grouping return lines is simple in concept; the complications are in planning the grouping pattern changes with growth from, say, one-for-one to one-for-nine. These complications are beyond the scope of this paper.

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Power Supplies

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(Manuscript received March 21, 1968)

This article describes the power supply system and the equipment developed to supply power to the repeaters and equalizers spaced along the coaxial cable. Power is supplied as a constant direct current over the center conductors of each pair of coaxial tubes in a series circuit consisting of two or four current regulated converters and the cable system consisting of the repeaters and equalizers. The system is designed for high reliability, long life, ease of maintenance and repair, and unattended operation.

I. INTRODUCTION

This article describes the power feed system and the equipment used to supply dc operating power to the repeaters and equalizers spaced about every two miles along the cable system. Included is a description of the integral protection and monitoring features which protect the cable system components from damage by excessive currents or voltages and which aid in diagnosis of equipment malfunctions.

II. POWER FEED SYSTEM

2.1 *Basic Arrangements*

The power feed system shown in Fig. 1 is used for each pair of coaxial tubes.[1] For systems shorter than 75 miles, one pair of power converters may be eliminated.

Power for the repeaters and equalizers is supplied over the coaxial tubes by circulating a constant direct current through the series circuit comprising the coaxial tube center conductors, the repeaters and equalizers, and the power converters. Each repeater or equalizer contains a voltage regulating diode which determines its voltage drop. With this arrangement the power delivered to each repeater or equalizer is independent of its location in the system.

No power current flows through the coaxial tube outer conductors.

These are grounded at each repeater point for personnel safety as one of the many precautions taken in the system design to prevent personnel contact with high voltages. The power separation filters shown in Fig. 1, and other filters located in each repeater or equalizer, direct the transmission signal and the power current through their respective paths.

Input power is obtained from a −24 volt battery system which provides operating power for all equipment in the station. The −24 volt battery system is of conventional design, using lead acid storage batteries maintained in a fully charged state by rectifiers operating from commercial ac power or from reserve ac gas turbine or engine alternator equipment installed in the station.

2.2 *Reliability*

Reliability refers to the ability of the system to provide continuous transmission capability. This capability is achieved by the use of redundant transmission facilities which are rapidly and automatically switched into service whenever working facilities fail. The extent of the redundant facilities required depends upon both the failure rate and the repair time of the equipment making up the system. In the L-4 system, redundancy is limited to the provision of a single complete Fig. 1 arrangement for protection of one to nine working systems.

The power converter is designed with sufficient margins against failures resulting from normal aging of components so that a malfunction almost certainly indicates catastrophic failure of a component. Passive components have been selected to provide a useful life of 20 years at their normal working stress levels. Low power semi-conductors are operated at stress levels expected to give failure rates approaching 0.001 percent per thousand hours. Power transistors are subjected to screening tests designed to yield devices which have failure rates not exceeding 0.01 percent per thousand hours under normal circuit operating conditions.

Self-contained alarm and monitor circuits facilitate diagnosis when trouble does occur. Faulty subassemblies can be readily replaced with spares, thereby permitting the equipment to be quickly restored to an operable condition. A low inherent failure rate combined with rapid repair capability makes this approach to system reliability practical.

2.3 *Basic Design Considerations*

In a dc power system such as this, it is necessary to guard against electrolytic corrosion resulting from sustained flow of dc current

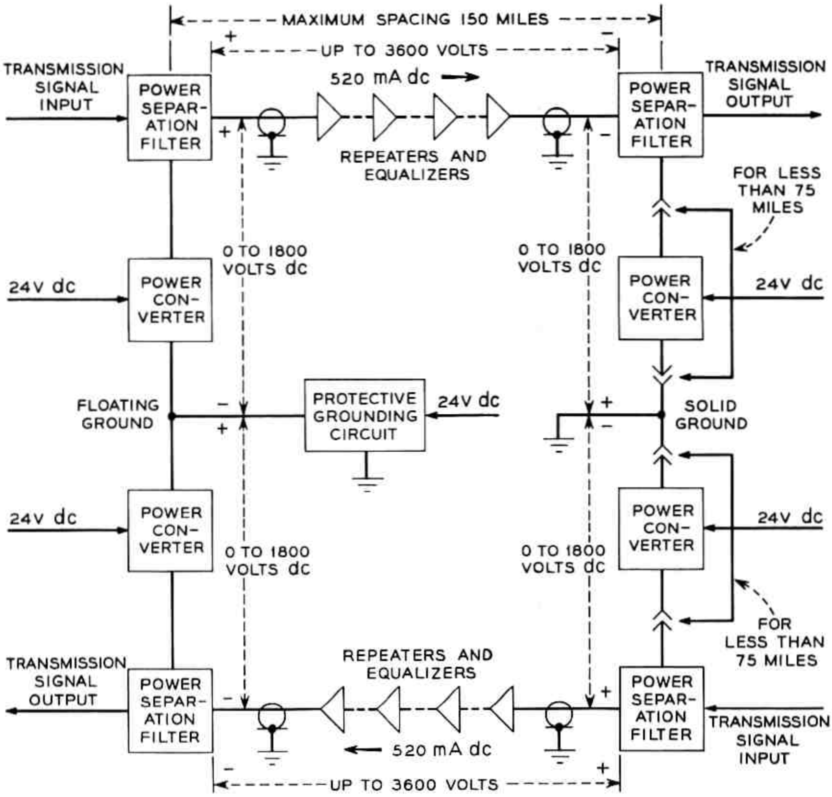


Fig. 1 — Power feed system block diagram.

through metal-earth interfaces. An ungrounded system would avoid the problem, but such a system has the disadvantage that the maximum spacing between power feed points can only be about one half that possible with a grounded system. The capability for maximum spacing between power feed points is desirable because the cost of the structure and facilities required at these points is a significant item in the overall system cost.

In an ungrounded system, the voltage to ground at any point around the loop depends upon leakage resistances which are not controllable in a working system. In extreme resistance unbalance, the voltage to ground at one of the power separation filters, for example, could be nearly equal to zero. In this case the maximum voltage point is at the other power separation filter located at the same end of the system.

The magnitude of this voltage is equal to the sum of the output voltages of two power converters and is twice what it would be if leakage resistance was balanced.

The corona threshold level of the coaxial tubes determines the safe upper working limit on voltage to ground. If leakage resistance is the only means for controlling voltage to ground, allowance must be made for the effects of resistance unbalance which, in the limit, requires a two-to-one margin. Since the required operating voltage is proportional to system length this two-to-one voltage margin necessitates a two-to-one reduction in the spacing between power feed points. If means are provided to limit the voltage excursions, then the spacing limitation can be eliminated.

After careful study, it was concluded that the best approach was to ground the system at one point as shown in Fig. 1. The required voltage limiting means is provided by the protective grounding circuit.

DC ground current can flow when the protective grounding circuit operates. This condition generates an alarm which serves to remind personnel that abnormal conditions exist and to stimulate action to return the system to the normal condition. The rate of corrosion when the floating ground is grounded is small; the total corrosion is not significant unless the abnormal condition exists continuously for several years.

III. PROTECTIVE GROUNDING CIRCUIT DESIGN

3.1 *Floating Ground Voltage*

The floating ground voltage consists of two components. The first is a low frequency (essentially dc) component resulting from variations in the individual converter output voltages, in the load voltage of the cable system and from earth potentials. The second is an ac component resulting from induction into the coaxial cable from nearby power lines and to transients arising from faults in the cable system.

When the grounding circuit operates, the resulting transient can momentarily affect transmission. It is desirable, therefore, that such grounding not occur except when there is real trouble in the system. Studies have shown that substantial ac voltages in the order of 800 volts can be induced, as a series (longitudinal) voltage, into the center conductors of the buried coaxial tubes when single phase faults occur on commercial power transmission lines which run parallel for distances as short as 15 miles to the cable route. The induced voltage is of similar magnitude and phase in each coaxial tube and therefore affects the instantaneous floating ground voltage. This induced voltage is troublesome only if its presence causes a malfunction such as insulation breakdown. The protective grounding circuit is designed to ignore this voltage unless it carries the instantaneous floating ground voltage and in turn the coaxial tube center conductor voltage beyond levels which will significantly reduce component life. A level of 2600-volts peak has been set as the maximum safe instantaneous voltage limit. This corresponds to a peak floating ground voltage of 800 volts. When the instantaneous floating ground voltage is carried above 800 volts it is necessary to follow the lesser of two evils and ground the floating ground. The resultant fluctuations in cable current caused by the induced voltage must then be tolerated.

In contrast to the maximum allowable short duration voltage of 2600 volts, is the maximum long term overvoltage which the system can withstand without jeopardizing system life. This voltage is about 2200 volts which corresponds to a floating ground voltage of 400 volts. If the low frequency component of the floating ground voltage exceeds 370 volts the protective grounding circuit responds and grounds the floating ground.

3.2 *Protective Grounding Circuit Description*

Figure 2 is a simplified schematic diagram of the protective grounding circuit. Cold cathode gas tubes, in a relaxation oscillator circuit, provide the low frequency threshold voltage sensing capability. Gas tubes were used because of their low leakage current at voltages below breakdown. A 20-millisecond time constant is provided by elements R1–C1 to make this part of the circuit respond only to the low frequency component of the floating ground voltage.

Protection for the ac component is provided by a carbon block protector with an 800-volt nominal breakdown voltage.

The output pulses from the relaxation oscillator are coupled through transformer (T1) to a transistor pulse stretching circuit which will operate the grounding relay (P) and alarm relay (A) with a single pulse from the relaxation oscillator. The alarm relay locks the circuit in the grounded state in addition to signaling the external alarm system. The ground and alarm are released by operation of the manual reset switch which opens the alarm relay lock-up path.

A long-life incandescent lamp, mounted adjacent to the gas tubes, provides illumination which results in rapid ionization of the gas tubes when the applied voltage exceeds their breakdown level. Loss of illumination can result in ionization times of up to a few seconds which is a tolerable but undesirable long-term operating condition.

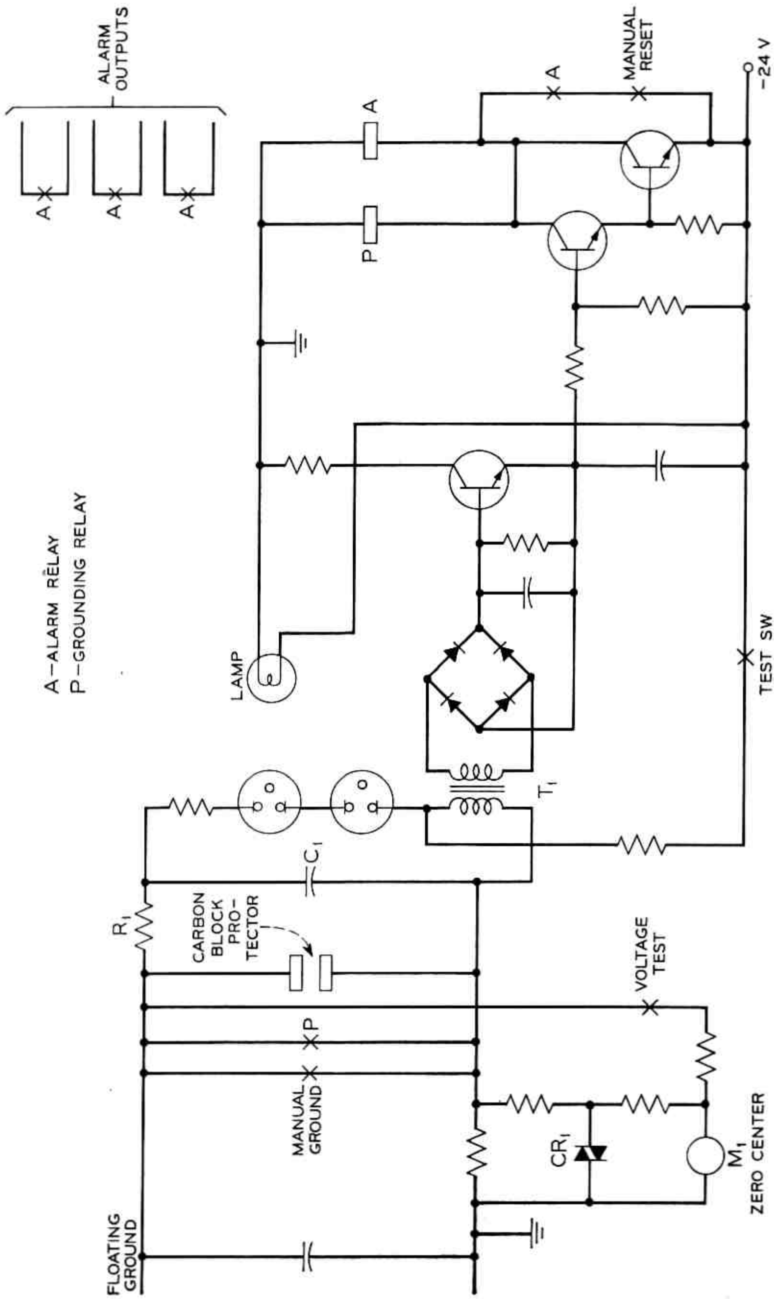


Fig. 2 — Ground circuit simplified schematic.

Two gas tubes that have a nominal breakdown voltage of 185 volts are connected in series to obtain the 370-volt low-frequency threshold voltage.

When the cable system is being energized or de-energized the floating ground is manually grounded. This permits power to be applied or removed on one coaxial tube at a time.

Meter M1 serves both as a milliammeter and a voltmeter. When energizing the system, a milliammeter null indicates that the currents in the two coaxial tubes are equal and the floating ground can be ungrounded without introducing power current transients. Likewise, when testing the ground circuit or when preparing to turn down power, if the floating ground voltage is zero, ground can be applied without introducing transients.

The pulse stretching circuit may be tested by operating the test switch. Before making this test the floating ground voltage is brought to zero by adjusting a power converter until a voltage null is indicated on the meter.

Diode CR1 serves to protect the meter against excessive current and permits a metering circuit having an expanded scale near null.

IV. POWER CONVERTER DESIGN

4.1 *General Considerations*

The power converter provides power to the cable system as well as providing a portion of the overall cable system automatic protection and alarm system. The latter function permits the cable system to be operated with a minimum of human supervision with the knowledge that the system will automatically protect itself from damage if abnormal conditions develop.

It was desirable, for both economic and technical reasons, to be able to tailor the power converter output capability to the requirements of the various cable lengths to be powered. Excessive output capability can cause system damage if the power converter regulator malfunctions and causes the power converter to deliver its maximum possible output into a short length system. This objective was readily achieved by the subdivision of the power conversion circuit into several identical modules or power stages and by arranging the circuit so that one or more power stages could be provided as needed to meet load requirements. A maximum of five modules are used, the number being dictated by the power handling capability of the available power transistors.

The voltage drops of the cable, the repeaters, and the equalizers are well controlled in manufacture but have only a small percent variation principally because of operation of auxiliaries such as the monitoring oscillators.[2] The load voltage sharing between the series connected power converters depends upon the magnitude and stability of their equivalent output resistances and the dc stability of their regulators. A lower bound on dc output resistance of 6000 ohms for a full length system is required to accommodate normal load voltage variations without exceeding the cable current deviation limits.

Table I contains a tabulation of the power converter electrical requirements and performance characteristics. The largest single variable to be accommodated is the input voltage variation. Open loop compensation is used to reduce the effects of this major variation and in

TABLE I — ELECTRICAL REQUIREMENTS AND PERFORMANCE CHARACTERISTICS

Converter input current	65 amperes maximum
Converter output current	520 milliamperes, working smoothly adjustable from zero
Converter input voltage	20–29 volts dc
Converter working output voltage range	40–1800 volts dc
Output current regulation and stability	±5.8% for combined variations in: temperature of: 40°–80°F input voltage: 20–29 volts output voltage: ±3.5%
AC component superimposed on dc output current	10 milliamperes rms maximum
AC component superimposed on dc input current	50 milliamperes rms maximum
Automatic shutdown limits	If output current decreases below 390 to 350 milliamperes or increases above 645 milliamperes after 1/2 to 1 second delay or if the output current decreases by more than 250 milliamperes in 1/2 second or fluctuates at an amplitude of 125 milliamperes at anywhere from a 4 Hz to a 2 kHz rate.
DC output resistance	1200 ohms per power stage 6000 ohms total

combination with a simple low-gain (3.5 dB) closed-loop regulator is sufficient to meet performance requirements.

4.2 *Power Stage Design*

Figure 3 shows a simplified schematic diagram of a power stage. Output control is achieved by controlling the pulse width of the "on" time of the power transistors which operate in the switching mode at a fixed repetition rate.[3] Pulse width is controllable from zero to virtually 100 percent duty cycle. Waveshape V_o is representative of the output from diode bridge CR5. This pulse train is filtered by the L1–C1 filter elements to produce the desired dc output.

Reliable operation of this circuit requires that simultaneous conduction of transistors Q4 and Q5 does not occur. The application of forward base current to an "on" going transistor is prevented until the off going transistor has fully turned "off." This requires that there be a time interval, defined as dead time, when no forward base current is applied to either transistor. This dead time must be greater than the difference between the actual circuit operating transistor turn-off and turn-on times.

Dead time is provided by the action of diodes CR3 and CR4. These diodes are designed to store a greater charge in response to a given forward current than the transistors. Consider current $ib4$, which for the moment is assumed to have the waveshape designated base drive current. When positive it passes in the forward direction through both the base-emitter junction of transistor Q4 and through diode CR4. When $ib4$ reverses polarity, at the beginning of a new half cycle of the base drive current, the charge stored in diode CR4 prevents reversal of its voltage until the stored charge has been removed. Until this occurs there is insufficient base voltage available to result in forward base current in transistor Q5.

The base drive current also passes through the base terminal of transistor Q4. This reverse polarity base current accelerates removal of the charge stored in the transistor reducing its turn-off time. When the transistor recovers, $ib4$ transfers to diode CR3. When diode CR4 recovers, $ib4$ transfers to the base of transistor Q5 and it turns "on." Thus the required dead time is provided. The process is similar for transistor Q5 and diode CR3.

Duty cycle control of transistors Q4 and Q5 is obtained by terminating forward base current prior to the end of each half cycle of the base drive current. Transformer T2 and the CR1 diode-transistor Q3

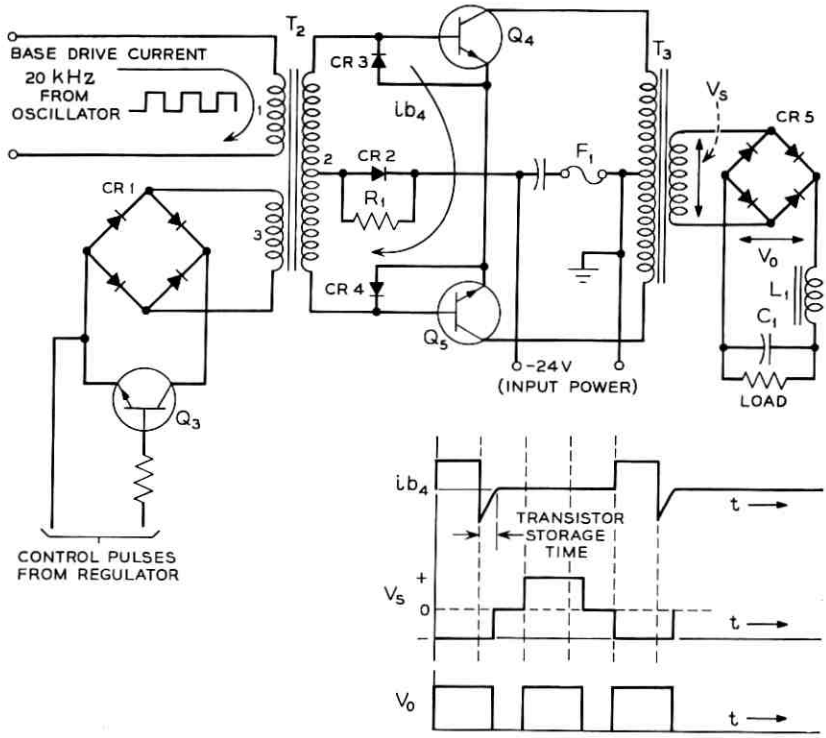


Fig. 3 — Power converter simplified schematic.

combination may be thought of as a current gate. When transistor Q3 is in the "off" state the gate is open and the base drive current passes through to the power transistors. When transistor Q3 is in the "on" state the gate is closed and base drive current is blocked.

To obtain satisfactory circuit operation it is necessary to minimize the time delay between the transistor Q3 control pulse and the power transistor Q4 or Q5 turn-off time. Power transistor storage time, the principal time delay, is minimized by taking further advantage of the charge stored in diodes CR3 and CR4. The transformed base drive current flowing through transistor Q4 and diode CR4, for example, results in stored charge in the forward biased junctions. Until this charge is removed these junctions have a definite open circuit voltage. When transistor Q3 turns on, the terminals of diode CR4 and the base and emitter terminals of transistor Q4 become parallel connected through the low impedance of winding 2. The resulting circulating current is in the reverse direction through the transistor base terminal. This accelerates the removal of the transistor stored charge. With this effect included current $ib4$ has the waveshape shown in Fig. 3. The decay in $ib4$ with time is caused by the increase in the transistor base-emitter impedance as the transistor stored charge is removed.

The time delay, resulting from the power transistor turn-off time, adds distortion between the control signal applied to transistor Q3 and the output pulse width. This is troublesome in applications such as this where the power stage functions as a pulse width modulator. Fortunately, as the output pulse width is narrowed, the finite power transistor response time results in less transistor stored charge making control down to zero output possible. This extreme control range is necessary to permit cable power turn up from zero.

Transistors Q4 and Q5 have specifications which limit the maximum in circuit turn-off time to 1.5 microseconds. Typically measured circuit turn-off time, at normal power output, is 1.0 microsecond.

Resistor R1 provides a path for collector-base junction saturation currents of transistors Q4 and Q5 during time intervals when transistor Q3 is "on" and no external base current is provided. Diode CR2 protects resistor R1 against excessive voltage if either transistor Q4 or Q5 develops an internal collector-to-base short circuit. The low impedance of the diode insures that this current will be large enough to blow the input protective fuse F1.

4.3 *Regulator Circuit Design*

To obtain open loop compensation against input voltage variations, the regulator circuit must develop a control signal for the power stage which results in generation of constant area pulses. This requires that the output pulse width V_o vary inversely with the magnitude of the input voltage. To obtain closed loop regulation it must be possible to vary, in response to a feedback signal, the area of the pulses held constant by the open loop compensating circuit.

The required control characteristic can readily be achieved by a magnetic amplifier. The magnetic amplifier contains two square hysteresis loop cores with windings arranged on these cores so that the magnetic flux swing in the cores can be controlled in response to currents and voltages impressed on the various windings.

Each core has an individual winding designated the gate winding. The two cores are then stacked and two additional windings wound around both cores. These windings are designated control windings. One of the control windings, further designated the bias winding, is current is in the reverse direction through the transistor base terminal. This accelerates the removal of the transistor stored charge. With this effect included current $ib4$ has the waveshape shown in Fig. 3. The decay in $ib4$ with time is caused by the increase in the transistor base-emitter impedance as the transistor stored charge is removed.

The time delay, resulting from the power transistor turn-off time, adds distortion between the control signal applied to transistor Q3 and the output pulse width. This is troublesome in applications such as this where the power stage functions as a pulse width modulator. Fortunately, as the output pulse width is narrowed, the finite power transistor response time results in less transistor stored charge making control down to zero output possible. This extreme control range is necessary to permit cable power turn up from zero.

Transistors Q4 and Q5 have specifications which limit the maximum in circuit turn-off time to 1.5 microseconds. Typically measured circuit turn-off time, at normal power output, is 1.0 microsecond.

Resistor R1 provides a path for collector-base junction saturation currents of transistors Q4 and Q5 during time intervals when transistor Q3 is "on" and no external base current is provided. Diode CR2 protects resistor R1 against excessive voltage if either transistor Q4 or Q5 develops an internal collector-to-base short circuit. The low impedance of the diode insures that this current will be large enough to blow the input protective fuse F1.

4.3 *Regulator Circuit Design*

To obtain open loop compensation against input voltage variations, the regulator circuit must develop a control signal for the power stage which results in generation of constant area pulses. This requires that the output pulse width V_o vary inversely with the magnitude of the input voltage. To obtain closed loop regulation it must be possible to vary, in response to a feedback signal, the area of the pulses held constant by the open loop compensating circuit.

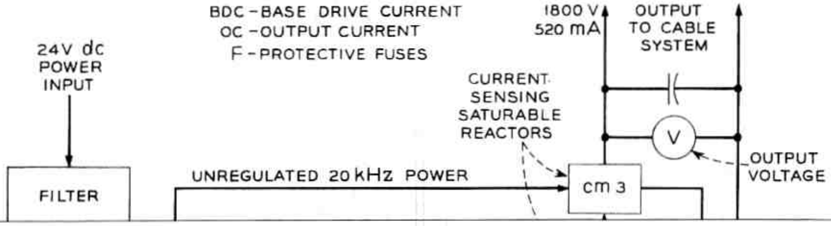
The required control characteristic can readily be achieved by a magnetic amplifier. The magnetic amplifier contains two square hysteresis loop cores with windings arranged on these cores so that the magnetic flux swing in the cores can be controlled in response to currents and voltages impressed on the various windings.

Each core has an individual winding designated the gate winding. The two cores are then stacked and two additional windings wound around both cores. These windings are designated control windings. One of the control windings, further designated the bias winding, is

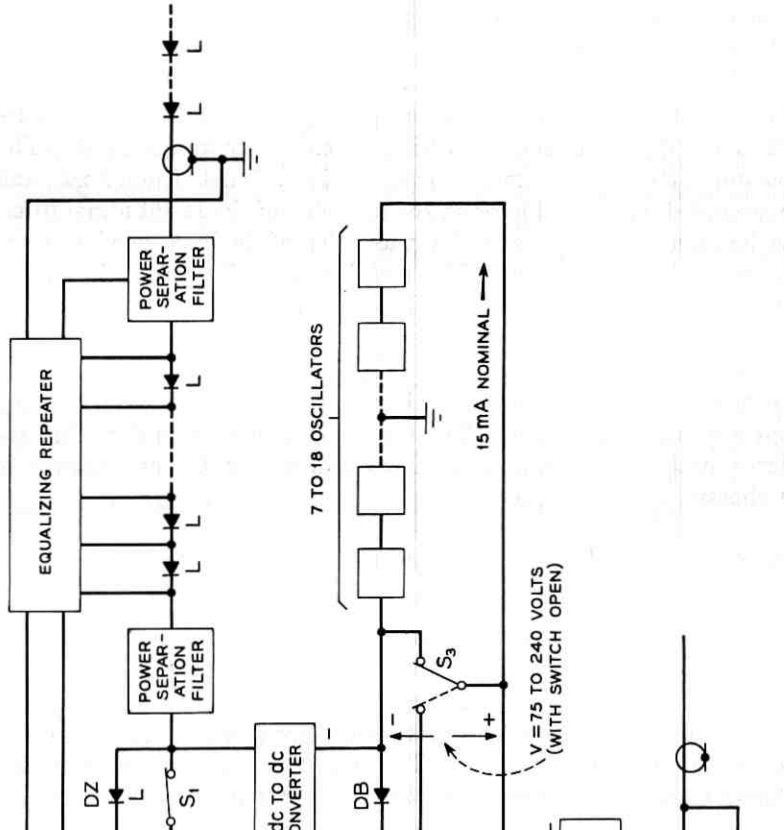
tem resulting in a virtually constant loop gain.

The required output resistance is, therefore, automatically obtained supplied a constant current and acts to bias or set the control characteristic for current in the other control winding at a convenient level and the required polarity.

In the absence of external gate winding voltage, the flux density



of 125 milliamperes or more in the frequency range of 4 Hz to 2 kHz.



- Monitoring oscillator power arrangements.

converter by diodes DB of Fig. 6. On-off control of these converters is provided from the transmission equipment acting through switches S1 or S2. A pair of converters, for redundancy, is provided at each equalizing repeater and is arranged to mount in manhole housings along with the transmission equipment.

Only one converter is energized at a time. Because of an upper limit (for personnel safety), of 120 volts to ground on the cable pairs feeding the monitoring oscillators, only 18 monitoring oscillators can be powered at one time. Switch S3 makes it possible to power a total of 36 monitoring oscillators in two groups of 18 each from one converter.

VI. PHYSICAL DESIGN OF THE POWER FEED CONVERTER

6.1 *Objectives*

The objectives of the physical design of the L-4 carrier power feed dc-dc converter power supply were to achieve compactness consistent with high voltage and heat dissipation requirements and to incorporate maintenance features to afford ease of rapid repair.

6.2 *General Description*

The complete converter power supply, seen in Fig. 7, is approximately 30 inches high, 21 inches wide, and 15 inches deep. The structure houses six large plug-in units, that is, up to five converter (power stage) units and one oscillator; and three smaller plug-in units, that is, the alarm sending unit, the alarm shutdown unit, and the regulating unit. Some of the plug-in units have been partially removed and are identified in the figure.

When less than five power stages are required, patch units, which provide circuit continuity through the connector and which prevent access to high voltages, are inserted in the unused positions. The converter power supply is suitable for use in a seven-foot-high standard power cabinets and nine-foot-high duct bays.

6.3 *Maintenance Feature Considerations*

Plug-in units, shown in Fig. 8, contain all of the active components, such as semiconductors and relays, to permit rapid repair of a power supply by replacement. Two extender designs for the plug-in units are available for maintenance and repair. Status lamps for alarm conditions, which remain lighted to retain a record of the fault cause if the unit is shut down, are furnished. Lamps, switches, and adjustments

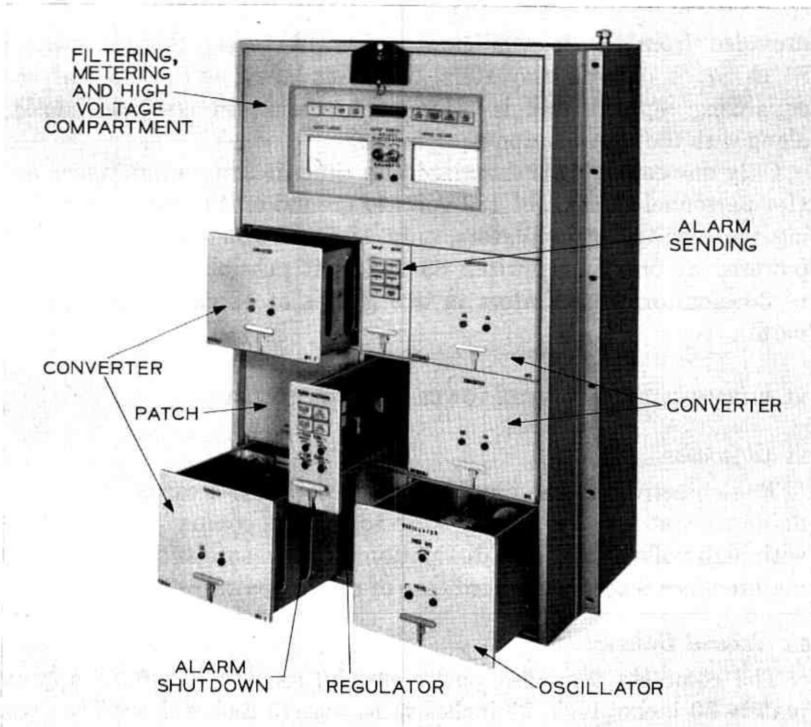


Fig. 7 — Power feed converter: Front view—plug-in units partially removed.

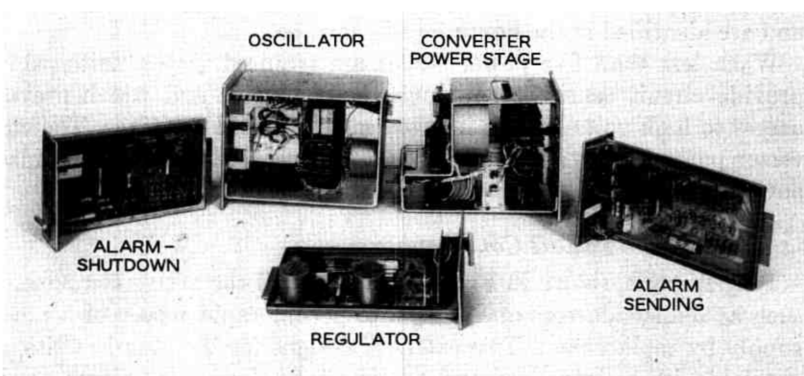


Fig. 8 — Converter plug-in packages.

for controlling and monitoring the power supply are located adjacent to the output meters to facilitate maintenance and adjustment.

6.4 *Personnel Protection Features*

The power supply's relatively high voltage of 1800 V requires adequate personnel protection features and mechanical design for long life operation. Certain parts of the equipment are protected by covers so that maintenance personnel are not exposed to hazardous potentials. The power supply series electrical interlock, described in Section 4.6, is an integral part of the personnel protection scheme. The converter (power stage) connector has interlock circuit pins which are shorter than those pins which carry high voltage, so that removing the converter plug-in will interrupt the interlock loop before the high voltage pins are disconnected.

6.5 *Detailed Mechanical Design of Plug-In Units and Extenders*

6.5.1 *Converter Power State and Oscillator*

The converter unit (power stage), shown in Fig. 7, is the basic building block module. The front panel latch, also shown, is used to eject the unit from the frame and to overcome the high contact forces of the high voltage connector at the rear of the unit. The power transistors are mounted on two heat sinks insulated from the chassis. Rectification is performed by a molded high voltage diode block assembly.

The 20 kHz driving frequency for the converter units is developed in the oscillator, also shown in Fig. 7. The oscillator has the same general profile as the converter unit. It uses the same type connector although different pins are used for the interlock loop, so that there will be no damage arising if a converter unit is plugged into an oscillator position or vice versa.

6.5.2 *Low Voltage Plug-In Units*

The three smaller plug-in units which may be seen in Fig. 7 are used for the basic regulating, alarm, and shutdown features of the power supply. These units have a die-cast frame and double-sided printed circuit boards with printed contact fingers to mate with the connectors mounted in the inner frame of the power supply.

6.5.3 *Extenders*

Two extender designs have been developed for this power supply for use when it may be necessary to repair the power supply or check various waveshapes in the field. The extender, for use with either the converter or oscillator unit (Fig. 9), is designed so that only the low voltage parts of the converter unit are accessible. These may be checked with proper maintenance tools when the extender is inserted in the power supply. It permits access to almost all of the components of the oscillator unit.

The extender unit for the three low voltage units is of much simpler design and incorporates a printed circuit board with a metal framework.

6.6 *Converter Main Chassis*

The power converter has a snap-in front cover which, when removed, exposes an inner hinged panel. Behind this panel are fixed components which include the 24 V battery input filter, a battery input fuse block with individual fuses for each power stage plug-in unit, and the battery input contactor.

Five combinations of voltmeters and associated multipliers are

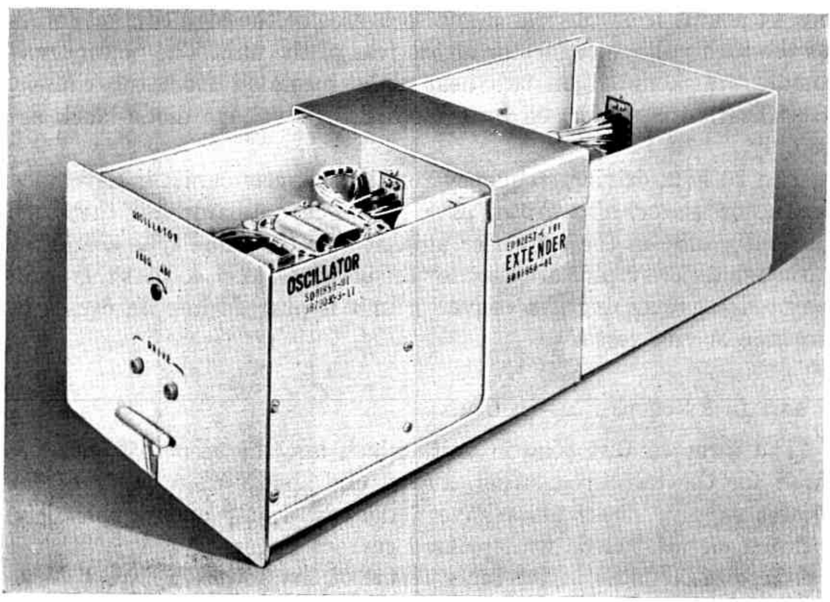


Fig. 9 — Extender for oscillator and converter (power stage) units—oscillator in position.

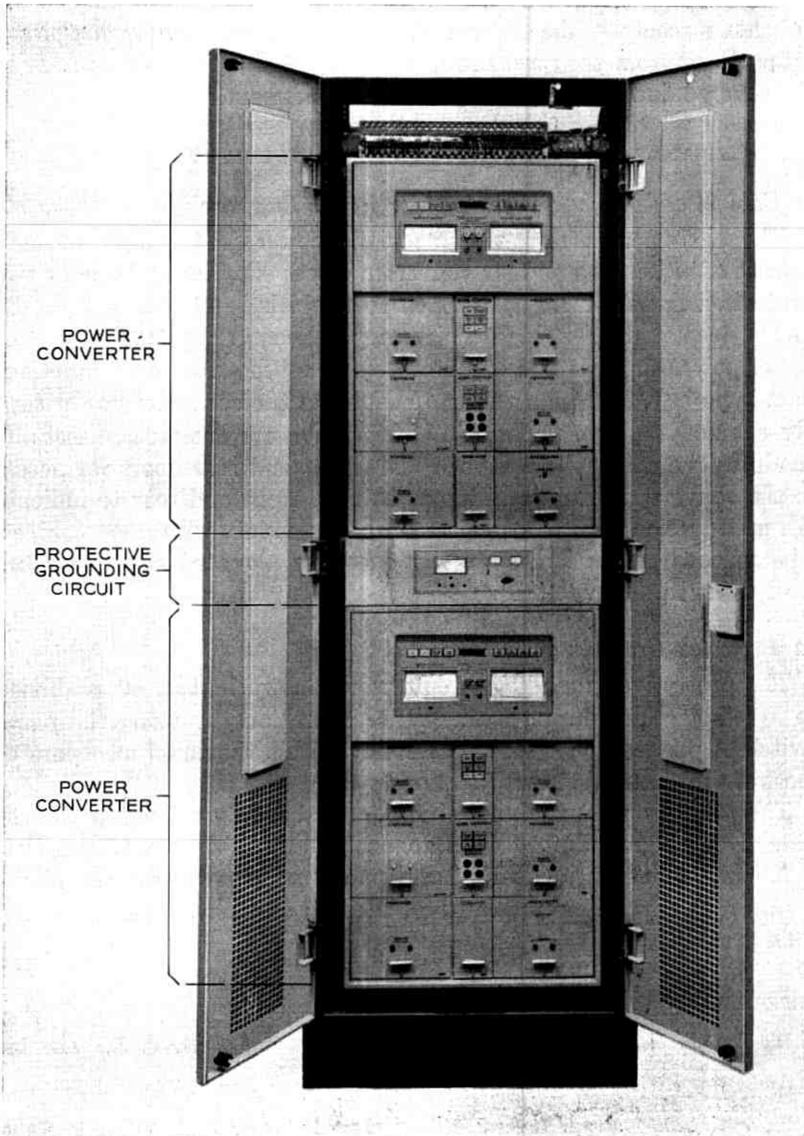


Fig. 10 — Power feed system equipment cabinet.

available consistent with the number of power stage units used. High voltage apparatus such as the output filters, the output current sensing saturable reactors, the voltmeter multipliers, and the high voltage output lead from the power supply to the coaxial cable system are in the covered, interlocked high voltage compartment.

VII. PHYSICAL DESIGN OF POWER FEED SYSTEM EQUIPMENT

7.1 *Cabinet and Bay Arrangements for Power Supplies*

The equipment arrangements for the power feed equipment may consist of either two or four converter power supplies and one or two protection panels enclosed in a cabinet which is 27 inches wide, 32 inches deep, and 7 feet high, which may be seen in Fig. 10, or two converter power supplies and one protector panel mounted on a nine-foot duct type bay. The high voltage leads from the converter power supply cabinets or bays to the L-4 carrier bays are enclosed in metallic conduit. The cabinet is equipped with front and rear doors for access to the converters and alarm lamps at both front and rear to indicate if a malfunction occurs in a converter power supply within the cabinet. The various status lamps on the duct-bay mounted converters are easily viewed.

7.2 *Protective Grounding Circuit Panel*

The protective grounding circuit panel shown in Fig. 10 is shifted to a solid ground during maintenance operations by using the panel switch. A printed wiring board assembly within the panel incorporates most of the essential circuit functions.

The solid ground panel, not shown, contains only a bus bar to provide a solid ground for series connected converter power supplies.

A supplementary jack and lamp panel, not shown, for the maintenance order wire circuit can be added to either the floating or the solid ground panel.

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Coaxial Cable and Apparatus

By G. H. DUVALL and L. M. RACKSON

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Impedance uniformity of the disk-insulated serrated-seam coaxial makes this Bell System standard a suitable transmission line for the L-4 system. As used in Coax 20, the nine in-service coaxial pairs and one reserve pair provide the capability of transmitting 32,400 simultaneous telephone conversations. To make the L-4 system a total communications network, support apparatus and enclosures which house and protect the electronic equipment from varied and rigorous environments were designed to complement the reliability and integrity of the Coax 20 cable.

I. INTRODUCTION

Greatly increasing requirements for long distance telephone circuits have made imperative the development of a new transmission system with greatly increased capacity. The transmission medium for the newly developed L-4 system consists of coaxial cable, manholes, repeater housings, and other required appurtenances. Critical to the system is the coaxial design. The uniformity of transmission characteristics over a very broad band of frequencies made the 0.375 inch disk-insulated serrated-seam coaxial the fundamental building block of the L-4 system.

The 0.375 inch coaxial cable shown in Fig. 1 has been standard in the Bell System for over 20 years; cables containing up to 12 coaxials were available until 1964. However, the increasing demand for telephone channels dictated the urgent development of Coax 20.

The transmission requirements established for the L-4 system necessitated placing repeater stations approximately every two miles along the cable route. At these locations, provisions must be made for the connection of electronic and associated support equipment to the cable.

Facilities at these locations include manholes, terminals, and sealed apparatus enclosures. The design and development of these facilities were directed at providing environmental, mechanical, and electrical protection for the electronic equipment at these sites. The natural hazards of total below-ground installations, that is, corrosion, water, humidity, induced earth currents, and lightning, require specific design treatment. In addition, hardening requirements to enhance the chances of system survival in case of nuclear attack, as well as the need for ready access to the electronic equipment for maintenance, represent additional criteria affecting design concepts. Reference 1 has a more complete description of system design parameters.

II. COAX 20, THE CABLE FOR L-4

2.1 *History of Serrated-Seam Coaxials*

Availability of the 0.375 inch disk-insulated serrated-seam coaxial shown in Fig. 1, was a major factor in the rapid development of a new system capable of transmitting over long distances a greater-than-ever cross section of simultaneous conversations. The coaxial has a 100-mil center conductor, air dielectric with polyethylene disk spacers, and a cylindrical outer conductor of 12-mil copper with interlocking serrated edges along a longitudinal butt seam. Two 6-mil steel tapes are helically applied over the copper to stabilize the structure and add electrical shielding.

The original version of (rubber) disk-insulated, serrated-seam coaxial has an electrical diameter (inside diameter of outer conductor) of 0.27 inch and saw its first commercial application in a 4-coaxial cable more than 25 years ago between Minneapolis, Minnesota, and Stevens Point, Wisconsin. As used with the L-1 carrier system, it provided a maximum capacity of 480 (later 600) two-way telephone conversations per pair of coaxials.

In 1946, a 0.375 inch coaxial, similar to the present design except for a 10-mil outer conductor, was installed in an 8-coaxial cable for an L-1 system between Dallas and Fort Worth, Texas. Its lower loss permitted an increase in distance between repeaters. This coaxial (with a 12-mil outer conductor) superseded the smaller design for future long distance installations.

A broadband L-3 commercial transmission system was introduced in 1953 on the improved coaxial, increasing the channel capacity per pair of coaxials to 1,860.[2,3] Overall cable capacity was increased again in 1959 when the 12-coaxial cable design was developed and put into manufacture for the transcontinental L-3 transmission system between Maryland and California.

As exploratory development established the feasibility of the new L-4 transmission system, no significant problems appeared in trans-

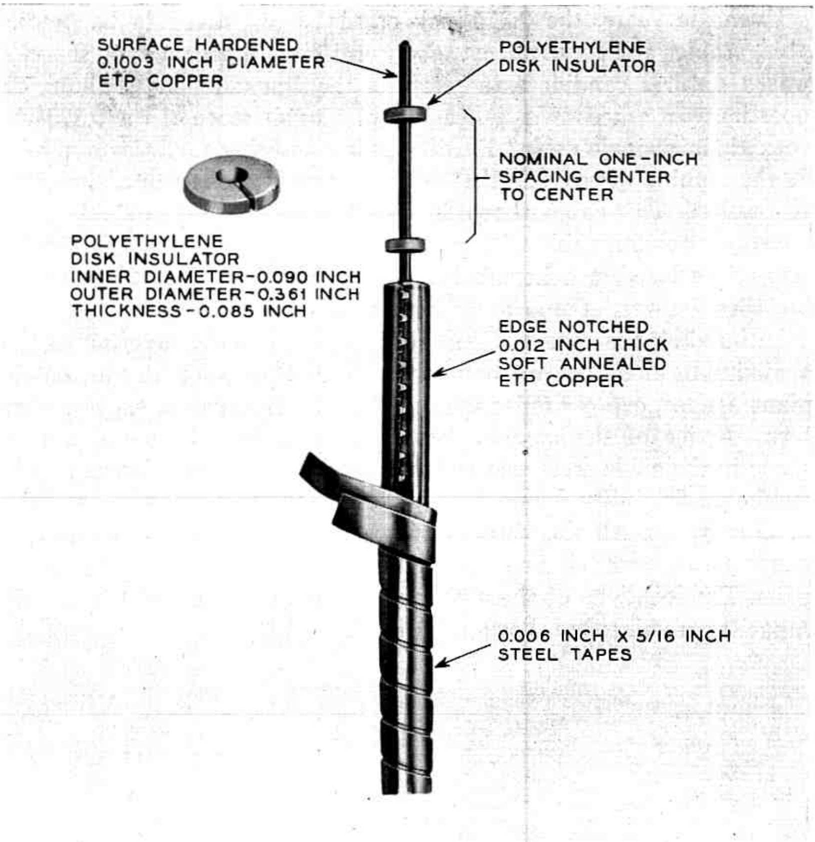

Fig. 1 — Air dielectric disk-insulated 0.375 inch serrated-seam coaxial.

mitting over the existing coaxial design. A field trial subsequently conducted in Ohio definitely established the feasibility of using the 0.375 inch coaxial for the new system.

2.2 *Development of Coax 20*

When forecasts of needed circuits indicated that the 9,300-channel cross section provided by the L-3 system on 12-coaxial cable was not adequate to keep pace with the anticipated demand for service, the American Telephone and Telegraph Company requested the urgent development of a larger facility. A facility was needed that could handle a growth of at least 2,500 voice channels per year.[4]

Over the years, the 0.375-inch coaxial had proved to be predictable, stable, and have an extremely uniform impedance; it therefore was a natural candidate to serve as the transmission medium. The question was asked: Was it feasible to provide more of the 0.375-inch coaxials in a single cable? Existing manufacturing facilities limited to 20 the number of coaxials that could be put into one cable, thus Coax 20 resulted. This cable, shown in Fig. 2 and described in Table I, is a two-layer coaxial cable with paper-insulated pairs and single conductors all enclosed in a Lepeth PJ sheath: The first Coax 20 cable was installed between Plano and Norway, Illinois, in 1964.

In building the core of Coax 20, it was originally determined that a minimum of 32 control pairs were needed for pilot alarms, outside plant alarms, express order wire, and so on. Because of expected long term service of the medium however, as many additional pairs as space in the cable cross section would permit were included to provide maximum flexibility for possible transmission systems to be developed in later years. All the wire circuits are 19 gauge except for four 16 gauge pairs, and the final design of the core consists of 52 control pairs. The locations of these paper-insulated circuits and their 1 kHz capacitance values are shown in Table I.

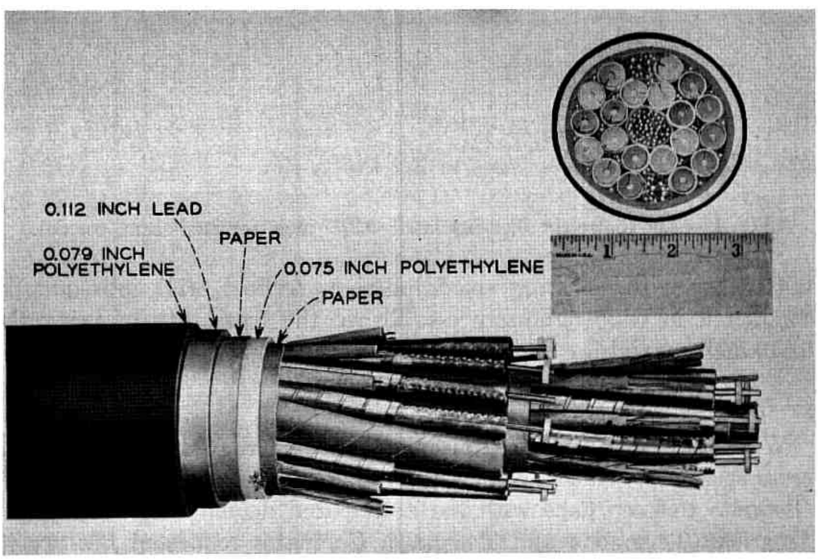

Fig. 2 — Coax 20: twenty 0.375 inch coaxials, forty-three pairs 19 gauge, four pairs 16 gauge, ten conductors 19 gauge, Lepeth PJ sheath.

TABLE I — DESCRIPTION OF COAX 20

Location	No.	Description
Center core	25	19 gauge pairs, 0.068 μF/mile
Inner coaxial layer	8	0.375 inch coaxials
	8	19 gauge interstice pairs, 0.078 μF/mile
Outer coaxial layer	12	0.375 inch coaxials
	2	units each (2 16 gauge pairs, 0.064 μF/mile) (5 19 gauge pairs, 0.065 μF/mile)
	10	19 gauge interstice conductors,* 0.089 μF/mile
Total	20	coaxials
	52	paired service circuits

* Capacitance to ground.

2.3 *Sheath*

Coax 20 is normally provided with a Lepeth PJ sheath which consists of an extruded polyethylene jacket over the core, a paper heat barrier, a lead sheath, a flooding of asphalt, and finally a black outer polyethylene jacket (Fig. 2). The reliability of this enclosure has been demonstrated by several years service of thousands of miles of coaxial cable. The lead provides the hermetic seal vital to maintaining the dryness of the core. The inner polyethylene layer provides high voltage protection between core and sheath. The outer polyethylene jacket provides protection against electrolytic corrosion of the lead. Layers of wire armor or gopher protection are added to the sheath when the cable is to be installed in areas where such added protection is needed.

2.4 *Manufacture*

The disk-insulated coaxial is made in a single operation on an automatic forming machine.[5] Immediately after forming, each coaxial is tested for corona and high voltage dielectric strength. A pulse echo test then measures the coaxial impedance uniformity and terminal impedance. The average of the terminal impedances for each coaxial is the basis for assigning the position of that coaxial in the final cable. The lowest impedance coaxial is assigned position 1, and so on. This scheme allows flexibility in the procedure of placing the cable and minimizes the mismatch that might occur as the cables are field spliced.

Stranding of coaxials into Coax 20 core takes place in two stages. The inner layer of eight coaxials is stranded over the core of 19 gauge pairs from eight positions of a two-bay, twelve-position floating carriage strander. Interstice wires and pairs (see Fig. 2) are stranded from fixed carriages attached to the back of one of the strander bays. After testing for impedance uniformity and high voltage requirements, the outer coaxials and associated paper-insulated circuits are stranded into position in a second operation. Because of differences in core diameter, the inner and outer coaxials are stranded using 20 and 36 inch lay lengths, respectively. This results in a helix or stranding take-up of 1.5 percent for all coaxials in the cable and provides for equal physical length of coaxials in both layers.

Each of the strander carriage positions is capable of holding sufficient coaxial to manufacture two lengths of Coax 20, each 1,750 feet long. The carriages are geared to impart a 37° backtwist to the coaxials as they are stranded into core. In effect, the steel tapes are tightened; this secures the longitudinal serrated seam, adds some rigidity to the structure, and provides a more uniform transmission line. After stranding, each core is tested for high voltage characteristics and impedance uniformity; then it is vacuum dried, sheathed, and retested. The finished cable is about 3 inches in diameter and can be shipped in maximum lengths of 1,750 feet. The average length has been 1,400 feet. When fully loaded, the shipping reels weigh about 8.5 tons each.

2.5 *Cable Characteristics*

2.5.1 *Attenuation*

The physical dimensions of the coaxial components shown in Fig. 1 and the parameters listed in Table II were used for computing the coaxial's attenuation at 55°F. This attenuation α was divided by the square root of the frequency, f, to permit a greatly expanded scale; the result is plotted in Fig. 3. The figure includes for comparison the attenuation-frequency characteristics measured on a one-half mile length of Coax 20 installed at the Chester, New Jersey, Bell Telephone Laboratories and corrected for stranding take-up.

2.5.2 *Impedance Uniformity-Echoes*

Echoes are divided into two catagories: internal and junction. Internal echoes are related to impedance discontinuities within a coaxial; junction echoes are related to the terminal impedance differences of coaxials joined in a field splice. The discontinuity associated with the splice components is discussed in Section 3.2.

TABLE II — PARAMETERS FOR THE 0.375-INCH COAXIAL

Parameter	Specification	
Surface hardened inner conductor	Diameter	0.1003 inch
	Conductivity*	100%
Polyethylene disks	Thickness	0.085 inch
	Spacing	1.00 inch
Annealed outer conductor	Thickness	0.012 inch
	Inside Diameter	0.369 inch
	Conductivity*	100.5%
Effective dielectric constant		1.098

* International annealed copper standard.

A measure of the worst internal echo in each coaxial is made at the factory using a 250 nanosecond raised-cosine pulse which contains a frequency spectrum to about 4 MHz. The averge of these data for 2,000 miles of coaxial cable was found to be 65.5 dB with a worst echo of 52 dB. A plot of these echoes expressed in dB is almost normally distributed as shown in Fig. 4.

The factory allocation of coaxials by impedance level minimizes the effects of impedance mismatch when coaxials in adjacent cables are field spliced. Factory-measured terminal impedance data was studied for 7,020 junctions, approximately the number needed for 2,000 miles of coaxial line; the calculated echoes resulting from mismatch were:

Condition	Echo	Number
Average	65.2 dB	—
Below	50 dB	25
Worst	48 dB	2

At 20 MHz, the junction echo resulting from the mismatch of coaxial

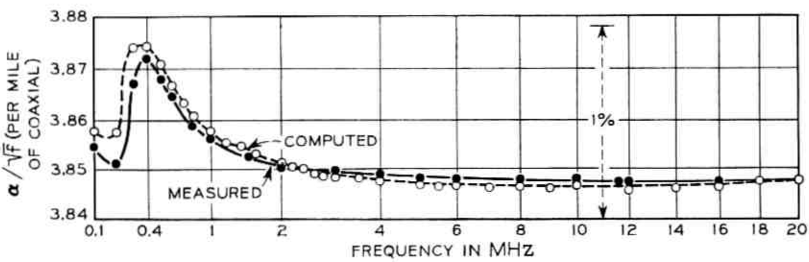

Fig. 3 — Attenuation divided by the square root of the frequency (per mile of coaxial). Derived from measurements on the installed Coax 20 at Chester, New Jersey (55°F).

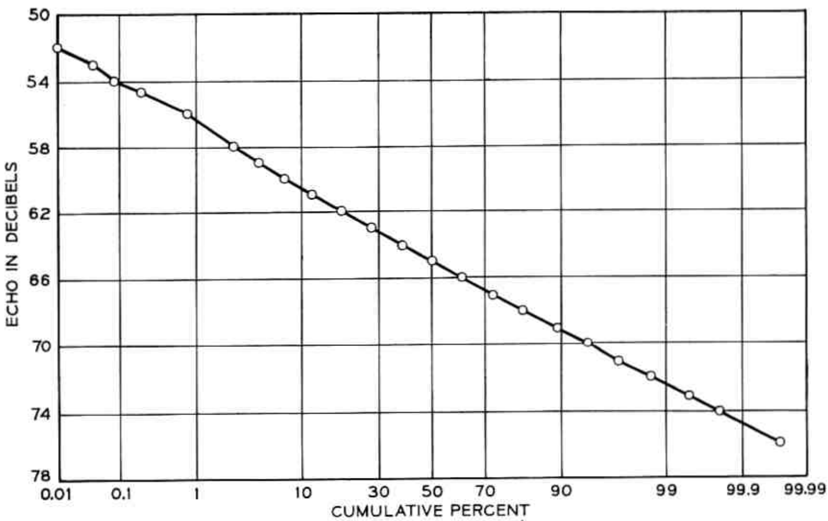

Fig. 4 — Distribution of the worst internal echo (out) for 7,520 readings. (Average = 65.62 dB; $\sigma = 3.48$)

impedance and splice components resolves to about 47 dB for the worst condition. When the impedances of the adjoining coaxials match, a junction echo of 65 dB caused by the splice is expected.

2.5.3 *Crosstalk*

The far end output-to-output crosstalk of adjacent coaxials in Coax 20 is plotted in Fig. 5. The plot shows the factory measured data of the inner and outer layer coaxials in 2,070 feet of cable (on its reel) and field data taken between repeater points (2 mile spacing) on normally installed Coax 20. Although it is difficult to assign causes for the difference in level, splicing and unit relaxation after installation are suspect. Additional comparative data taken at the Chester, New Jersey, and Baltimore, Maryland, Bell Laboratories also are shown.

2.5.4 *Structural Return Loss*

Shortly after Coax 20 went into commercial production, periodic deformation was observed in the outer coaxial layer. The deformation resulted from the cable pressing against guides in the factory process. Each coaxial presented itself briefly to the guide surface every 36 inches by virtue of its helix or stranding lay length. These irregularities escaped detection by the 250 nanosecond raised-cosine test but their presence was readily detected when a "structural return loss" test was instituted. For this test, the frequency of the input signal to the cable is swept from 8 to 220 MHz and the reflections are monitored. This test is extremely sensitive to periodic irregularities because the vector sum of reflections from regularly spaced deformations add inphase at the frequency for which the spacing is a half wavelength. At this frequency, not only is return loss maximum, but also a peak in the attenuation characteristic will occur.

The following expression can be used to predict at which frequency a structural return loss spike will appear for a given spacing of periodic deformation (conversely, the spacing can be determined if the frequency is known):

$$f = \frac{V}{2(\epsilon)^{\frac{1}{2}} S} W(1 - TU),$$

where

$$\begin{aligned} V \quad &= \text{velocity of light in air (feet/second)}, \\ \epsilon \quad &= \text{relative dielectric constant}, \\ W \quad &\approx \left\{1 - \frac{\text{internal inductance}}{2\text{ (space inductance)}}\right\}, \\ f \quad &= \text{frequency in MHz}, \\ S \quad &= \text{spacing between irregularities (feet)}, \\ TU &= \text{stranding take-up (1.5 per cent in Coax 20)}. \end{aligned}$$

For Coax 20 a spike will occur around 156 MHz which is the half wavelength frequency for the 36 inch stranding lay of the outer coaxial layer. As improved testing techniques become available, the structural return loss test frequency is being expanded to detect the impedance irregularities associated with the inner layer coaxial lay length. For this lay, a spike can be observed at about 285 MHz. Although these frequencies are well above the L-4 band, the coaxial structure irregularities are being corrected so that Coax 20 may be suitable for future analog as well as high-speed digital transmission systems.

III. INSTALLATION

3.1 *Placing*

Protection of the medium against some of the effects of a nuclear blast is provided by a "hardened" installation. Basically this requires a subterranean system with shield and shock resistant features built into the support apparatus and housings. Coax 20 is usually buried

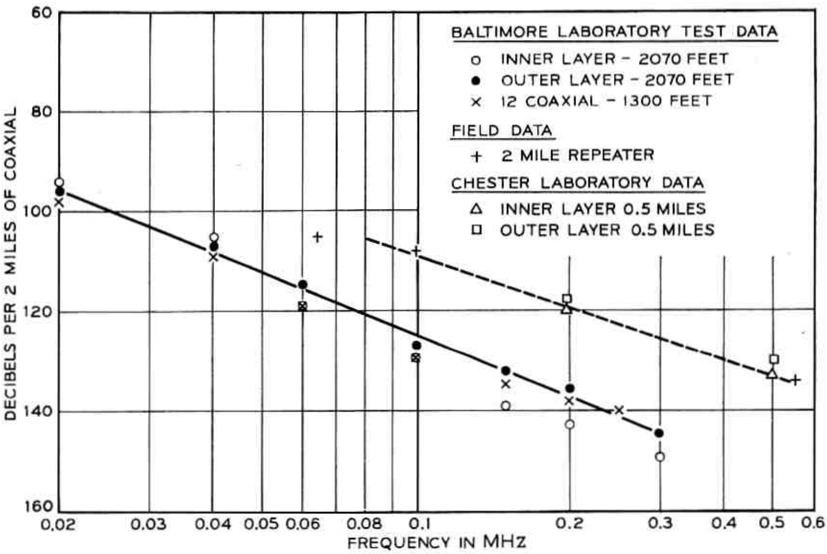

Fig. 5 — Far end crosstalk of adjacent coaxials for Coax 20.

a minimum of four feet in a trench along routes which avoid populated or strategic areas. Shield wires placed in the cable trench supplement the lead sheath conductivity. The trench digging, cable laying, and simultaneous backfilling and positioning 6 AWG copper shield wires about two feet above the cable are shown in Figs. 6, 7, and 8. Compared with coaxial cables installed by earlier methods, which did not specify shield wires or minimum four-foot depth, limited information indicates that the failure rate per route-mile for hardened cables may have been reduced by a factor of three.

3.2 *Field Splicing*

Figure 9 shows a typical field splice of a single coaxial and associated parts. The center conductors of adjacent coaxials are crimped into the S-100 sleeve; the outer conductor and steel tapes are compressed between the steel bushings and copper sleeves. Continuity is provided through the G-375 sleeve, which has its center section expanded to maintain, as closely as possible, a 75 ohm impedance through the whole splice. Based on measured components of the resistance and reactance in a splice, the total calculated return loss is 65 dB at 20 MHz. Return loss versus frequency is plotted in Fig. 10.

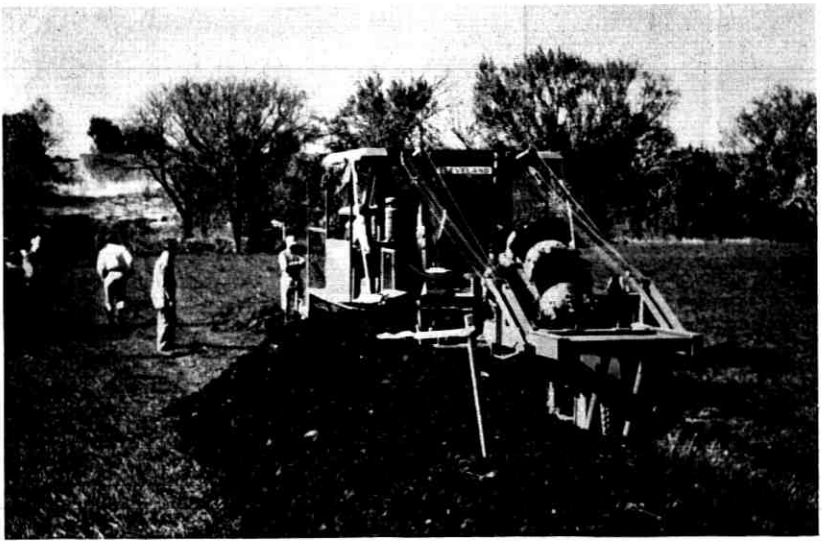

Fig. 6 — Trenching: four foot depth.

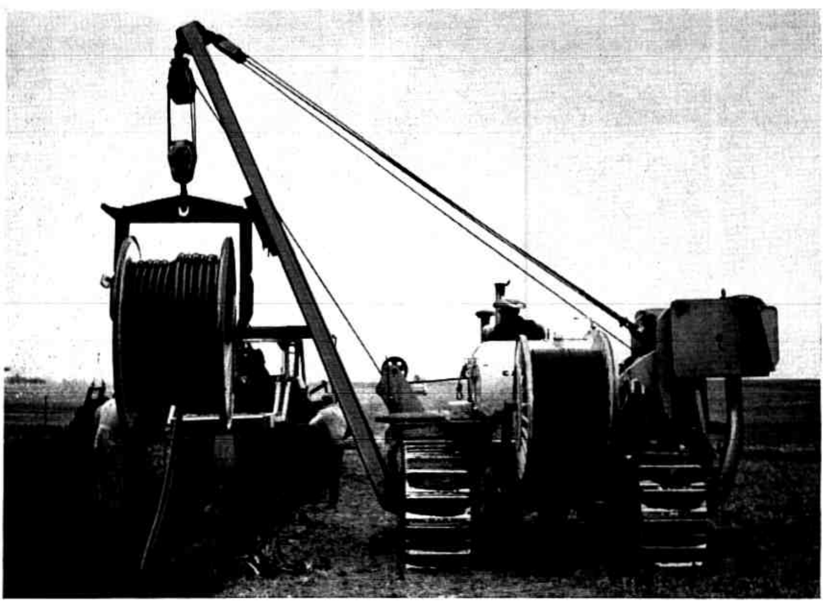

Fig. 7 — Laying Coax 20.

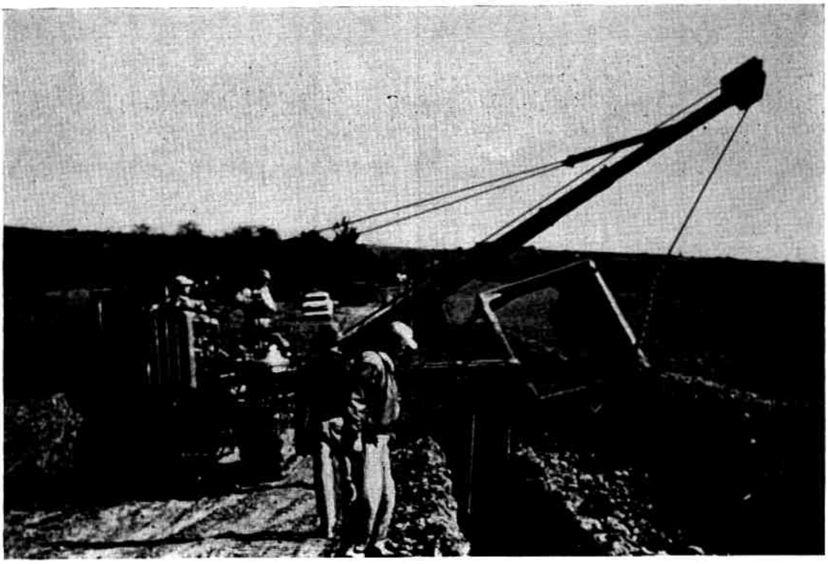

Fig. 8 — Backfilling and placing of shield wires.

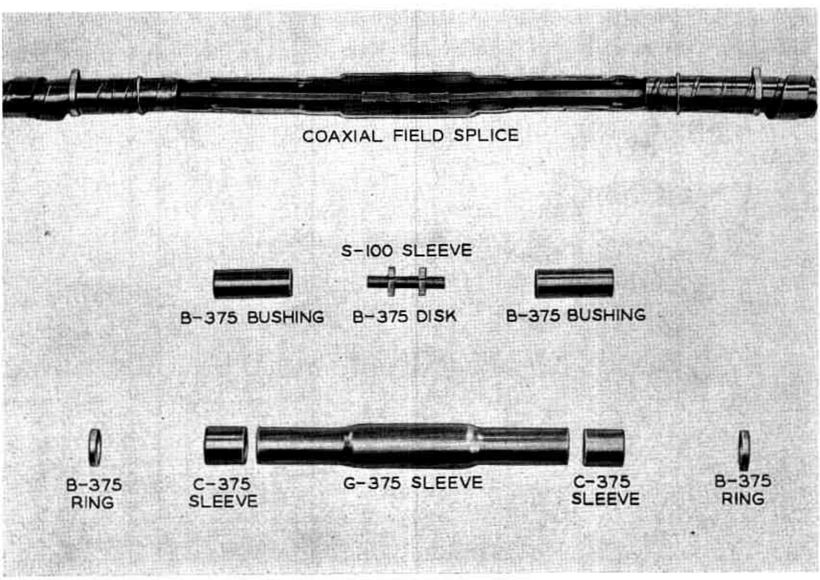

Fig. 9 — Coaxial field splice.

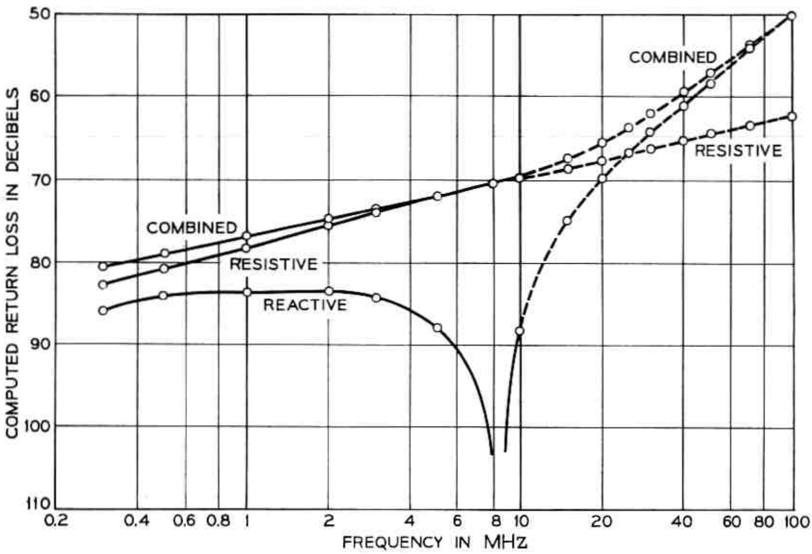

Fig. 10 — Return loss versus frequency for the G-375 coaxial field splice.

Normally, seven reels of cable are placed between adjacent repeater stations. The cables are spliced so that coaxials of similar impedance are joined together. The coaxial positions are identified by the color code of the paper-insulated interstice circuits. Figure 11 shows a typical "basket" formed by the spliced coaxials.

IV. APPARATUS FACILITIES

A typical equalizing cable section, shown schematically in Fig. 12, comprises one equalizing repeater station, three regulating repeater stations, and twenty basic repeater stations in which a multitude of additional support apparatus is required for the repeater to function properly.

The signal through the main coaxial cable is transmitted to the electronic equipment in the 471A1 apparatus case by the 66A1-4 cable terminal. As can be seen in the basic station schematic, Fig. 13, ten of these four unit coaxial terminals are connected to the ports of five apparatus cases, each capable of housing four basic repeaters and the monitoring oscillator. Coaxial patch cords carry the signal from the faceplate of the cable terminal to the repeaters.

The two 52-pair stub cables in the 472A1 apparatus case, which

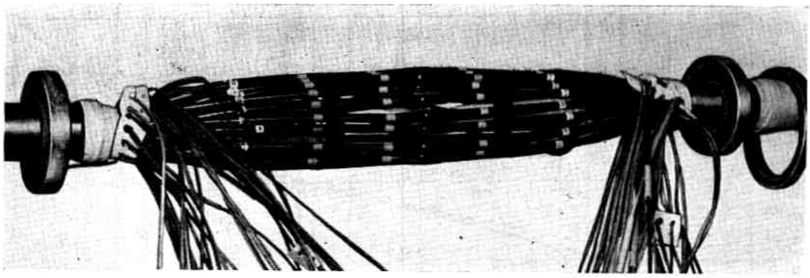

Fig. 11 — Basket formed of spliced coaxials. A steel sleeve is placed over the splice and solder wiped to lead sheath through lead disks and rings.

serves as a cross-connect facility, are spliced to the copper pairs in the main cable. The 26-pair stub cable feeds through the auxiliary splice to the 100A1-4 cable terminals attached to the 471A1 apparatus case to provide power for the monitoring oscillator and to the 100B1-4 cable terminals located in the collar of the manhole to provide voice facilities for the craftsman.

The schematic diagrams for the regulating and equalizing stations are shown in Figs. 14 and 15, respectively. The physical connections and function of the apparatus in these manholes are similar to the basic station, except for the amount of component apparatus required.

The following discussion illustrates the design philosophy behind some of the more complex items.

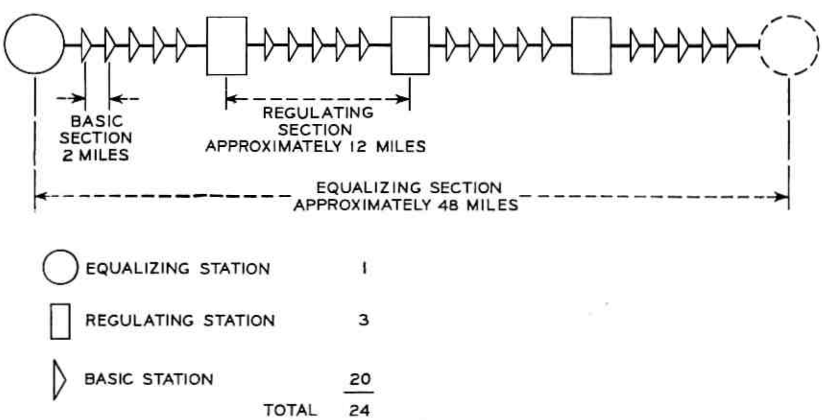

Fig. 12 — Equalizing section schematic.

4.1 *Manholes*

L-4 system requirements for a hardened installation together with limitations on space and economic considerations, dictated the design of special precast, segmented reinforced concrete manholes for the basic and regulating stations. The size required to accommodate all of the components associated with the equalizing station ruled out a precast design; therefore, a cast-in-place manhole was used.

The basic manhole, composed of three interlocking sections plus a collar, measures 10 feet long by 6 feet wide by 6 feet 6 inches high inside, and weighs 42,600 pounds. The regulating type, illustrated in Fig. 16, has four interlocking sections plus a collar, measures 12 feet long by 6 feet wide by 8 feet high inside, and weighs 58,400 pounds. Both manholes can withstand an overpressure of 100 pounds per square inch. Unistrut inserts are cast into the interior wall sections to support cable and equipment; inserts to facilitate handling and lifting are provided on the outer surfaces. The precast structures can be installed around an existing buried cable, existing conduit encased cable, or can be used on new cable construction.

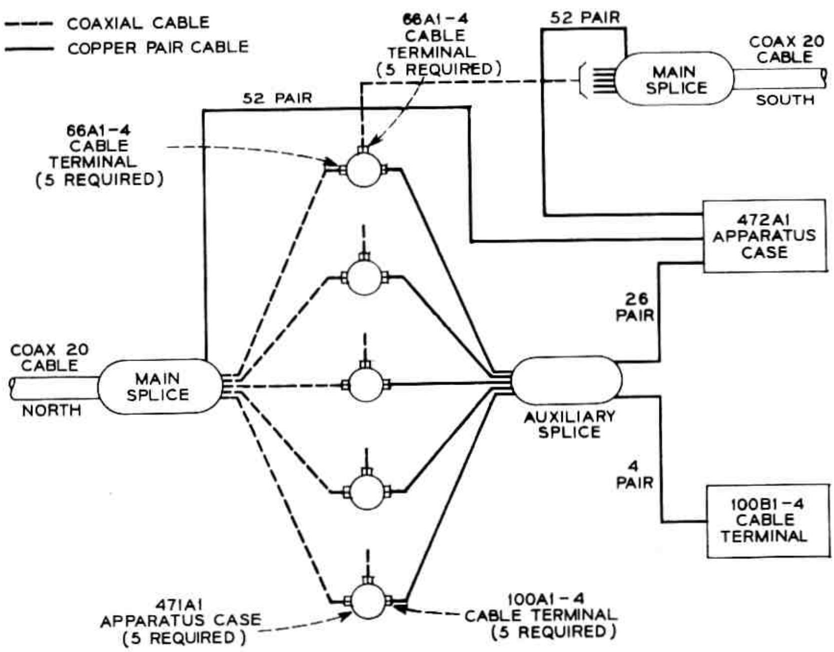

Fig. 13 — L-4 basic station schematic.

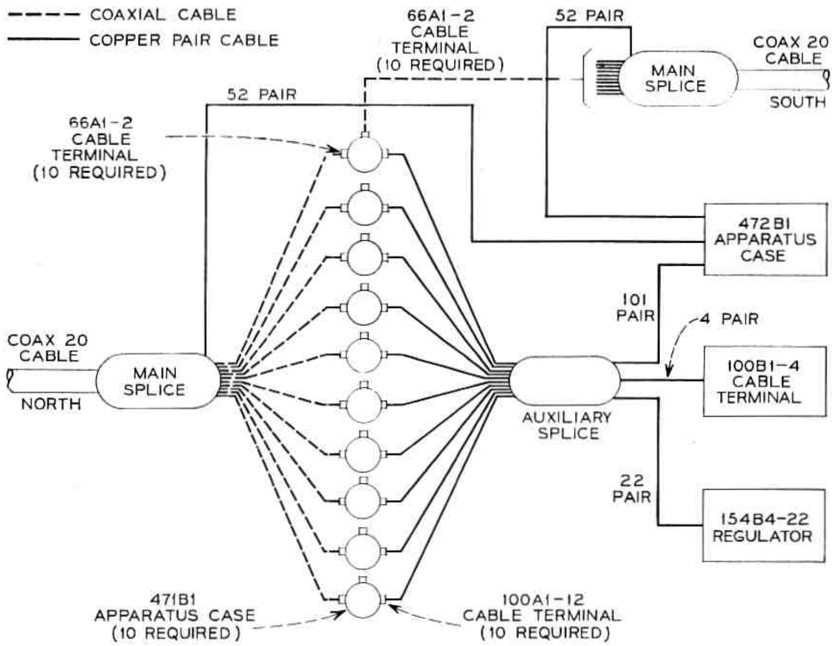

Fig. 14 — L-4 regulating station schematic.

The cast-in-place equalizing manhole is capable of withstanding a 50 pounds per square inch overpressure and measures 24 feet long by 12 feet wide by 8 feet high, inside. There is a unistrut framework down the center of the manhole for mounting auxiliary apparatus and cable.

4.2 *Apparatus Enclosures*

4.2.1 *Repeater Apparatus Case*

Environmental protection for the repeaters is provided by the sealed apparatus case shown in Fig. 17. These cases are available with internal chassis designs to accommodate one equalizing, or two regulating, or four basic repeaters. However, at equalizing points, each repeater and the associated control equipment requires two apparatus cases mechanically and electrically interconnected. The assembled units are about 16 inches in diameter, $22\frac{3}{4}$ inches high, and weigh 150 pounds.

Galvanic action between dissimilar metals, which is always present in below ground environments, demands the utmost care in choosing

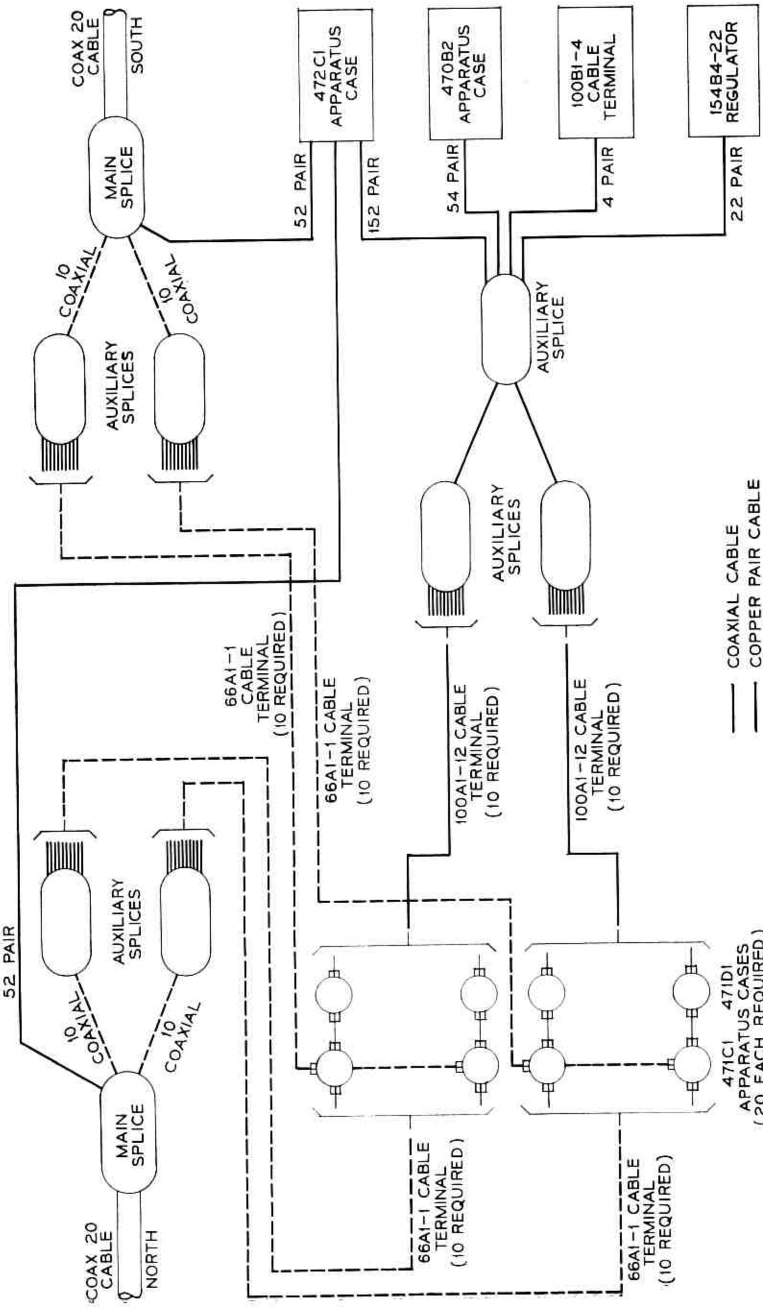

Fig. 15 — L-4 equalizing station schematic.

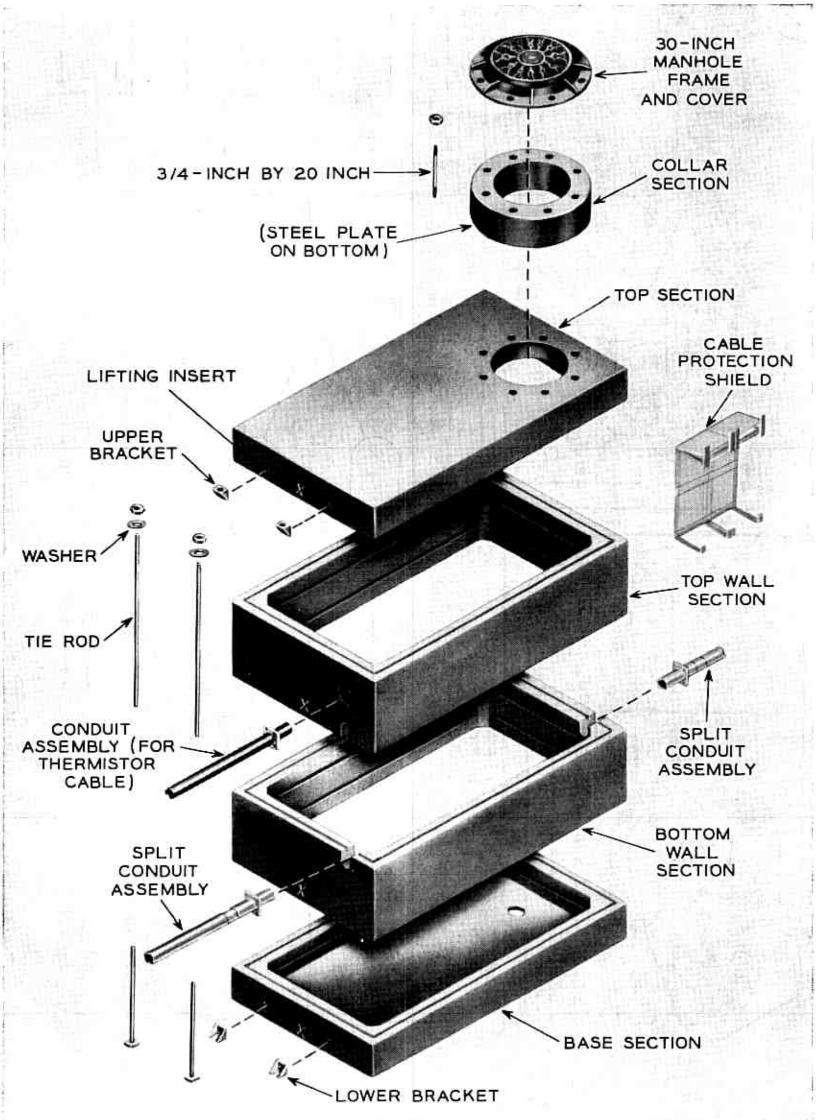

Fig. 16 — G manhole: exploded view.

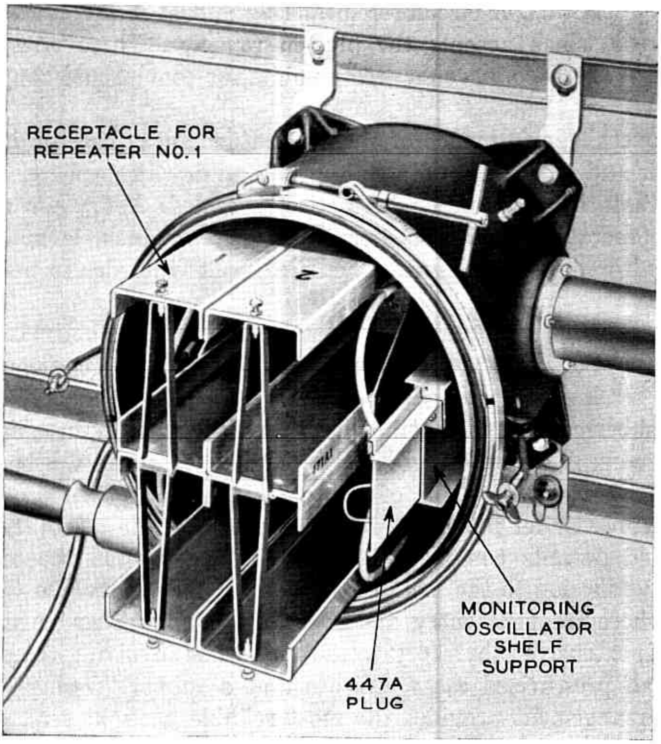

Fig. 17 — Basic repeater apparatus case with cover removed, mounted on the manhole wall.

the best available materials and coatings to provide long life for the apparatus housings. The choice of galvanized cast iron, where the casting process was most applicable, and fabricated galvanized steel is the result of many years of laboratory testing and evaluation to determine the best corrosion resistant material for underground use. Data on cast iron housings with a minimum wall thickness of 5/32 inch and a zinc coating of 2 ounces per square foot indicate a life expectancy of 20 years minimum in highly corrosive environments and more than 40 years in milder environments. These metals are also compatible with such other materials used in underground plant as the lead sheath on the cable and manhole racks.

The outer housing consists of a galvanized cast iron base with a 5/32 inch minimum wall thickness and a removable galvanized steel cover. Four reinforced pads are provided outside the base at the bottom, for mounting. A circumferential ring is cast inside as a seat for the chassis which accepts the plug-in repeater. Three openings on the side of the base serve as access points for coaxial and copper pair terminals.

The overall performance of the system demands that the inside of the cable and the repeaters be in a clean and dry atmosphere. This requires that the network be hermetically sealed. A dry air pressure system, operating at 6 to 9 pounds per square inch, is maintained and monitored to signal sheath and seal faults in order to permit repair before water enters.

The sealing technique involves two approaches, both based on experience. For apparatus where field assembly is a one-time operation, B sealing compound is used. This is a pliable, tacky, uncured butyl rubber compound that has been used successfully in splice cases for 15 years. A groove is provided around the ports in the base casting and on the face plate of the cable terminals to accept the B sealing cord. Under moderate clamping pressure, the pliability of the compound permits compact filling of the groove and the tackiness provides adhesion to the metallic surfaces of the groove. On the other hand, where frequent entry and resealing of apparatus housings are required, extensive laboratory testing programs involving temperature cycling of pressurized cases indicate that a rubber "O-ring-V-band" clamp arrangement provides the most reliable gastight seal.

An important feature of this type seal is that a load need be applied at only one point around the periphery to effect a seal. As tension is applied to the T-bolt, the V-band embraces the flanges of the cover and base for the full 360 degrees, thereby assuring equal distribution of pressure on the O-ring. The geometry of this assembly and the elastic energy stored in the V-band clamp provides the compliance needed to maintain positive pressure on the O-ring to assure its reliability. Figure 18 illustrates the V-band-O-ring principle.

Two slotted tabs are incorporated on the cover to accept an eyebolt and wing-nut arrangement and to provide a safety feature which prevents the cover from "blowing off" during disassembly in the event the pressure in the case has not been released.

Internally, the apparatus case has a removable fabricated aluminum chassis into which repeaters and associated control equipment are inserted. The receptacles are sized to provide a toleranced fit for the easy insertion of each type repeater. Although the outer wall of the receptacles "float" to aid initial engagement, a locking bar arrangement provided at the top prevents plug-in units from disengaging and re-

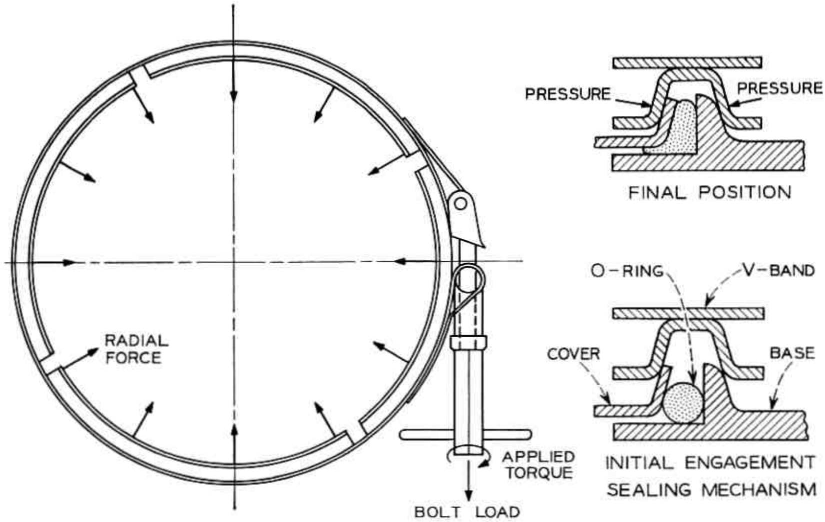

Fig. 18 — The O-ring-V-band sealing mechanism.

sults in a more positive and intimate contact between the surfaces of the repeater and the receptacle, enhancing heat dissipation. The eight slotted holes in the bottom plate serve a twofold purpose: they lock the chassis in the base and provide as much intimate contact between the base and the chassis as possible for maximum heat transfer.

Each chassis incorporates a network of coaxial cords and copper pairs terminated in connectors which provide an electrical link to supplementary apparatus. Short length coaxial patch cords are also provided to facilitate shorting the input side to the output side when required. The two chassis for the equalizing station are somewhat unusual in that additional connectors and two cords are required for some of the copper pairs in order to interconnect the two chassis electrically. To facilitate the mechanical interconnection between the two apparatus cases, a galvanized cast iron through pipe has been provided. The typical chassis for the basic apparatus case is shown in Fig. 19.

The temperature sensitive characteristics of the repeaters make it imperative to dissipate the heat generated by their operation. This required explicit design features in the apparatus cases to provide effective heat transfer paths. Power ratings for the basic, regulating, and equalizing repeaters were established as 13.8, 32.3, and 80.3 watts, respectively. However, the number of repeaters in the apparatus hous-

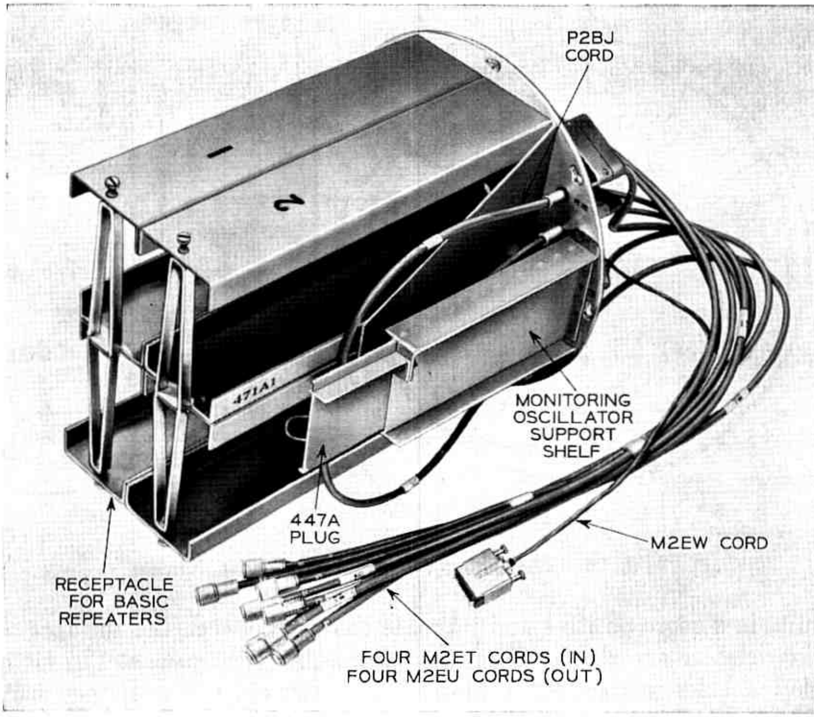

Fig. 19 — Chassis for basic repeaters.

ing varied according to type, that is, four repeaters in the basic housing and two repeaters in the regulating housing. This meant that the total power level was about 55 watts for the basic apparatus case and 65 watts for the regulating case. The equalizing repeater presented a somewhat different problem in that two apparatus cases were required to accommodate all of the plug-in equipment associated with the repeater. The power breakdown for the two cases was established as 60 watts for the transmitting case and 20 watts for the control case.

Initial heat tests utilized unpainted apparatus cases which had four internal lugs, on which the chassis seated, and four fastening screws. Likewise, the repeater units were unpainted. Additional tests indicated that an internal circumferential ring around the cast base, 8 fastening screws, and continuous welded joints between the upright structure and the mounting plate of the chassis improved the heat transfer mechanism. To further enhance this mechanism, an organic finish, black paint, was applied to the inner and outer walls of the apparatus case and cover and to the cover of the repeaters. A 23 percent improvement in the heat transfer path was realized as shown in Fig. 20.

Extensive laboratory shock tests were conducted at the Whippany, New Jersey, Bell Telephone Laboratories. A drop table simulated shock loads up to 250 g. The repeater apparatus case was tested in 3 planes and survived shock loads of over 200 g.

4.2.2 *Cross Connect Facility*

The 472 type apparatus case provides cross-connection and electrical protection features for the system and is similar to the repeater apparatus case in that an O-ring and V-band clamping arrangement is used to effect a gastight seal between a cylindrical cover and plate. The assembled case is approximately 11½ inches in diameter, 40½ inches long, and weighs approximately 125 pounds.

As shown in Fig. 21, a welded H-frame, which gives the desired strength and rigidity to meet the hardening requirements, is utilized to accommodate connecting blocks. Three 22 gauge PE-PVC (polyethylene-polyvinyl chloride) insulated Lepeth sheath stub cables are wired internally to the connecting blocks in each apparatus case. The

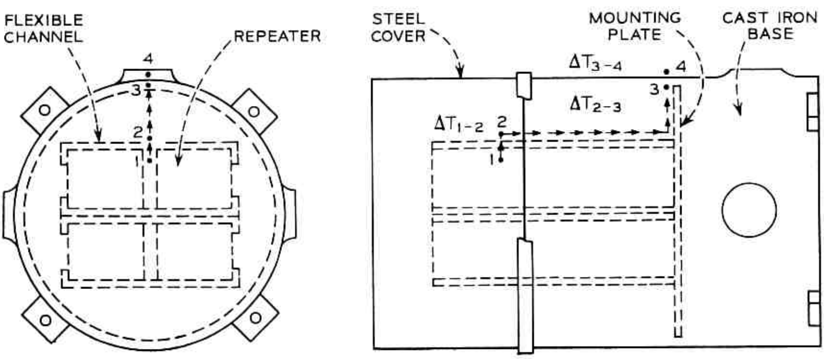

Fig. 20 — Temperature differentials in the L-4 basic repeater apparatus case. The length of the heat transfer path is 15 inches.

	ΔT — °F			
	ΔT_{1-2}	ΔT_{2-3}	ΔT_{3-4}	ΔT_{1-4}
Unpainted	17	17	6	40
Painted	13	15	3	31

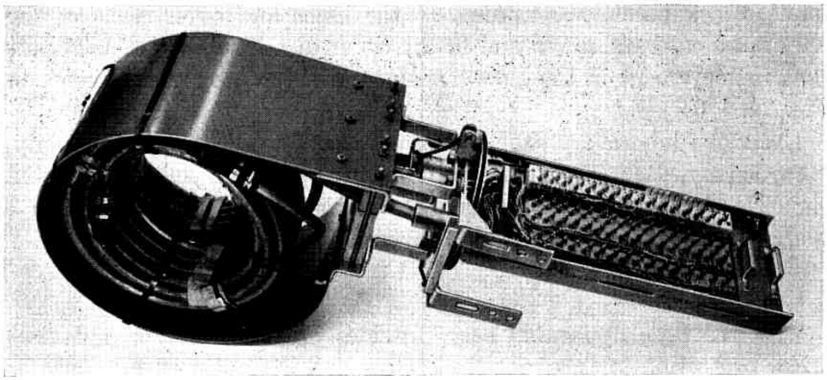

Fig. 21 — Cover plate and internal chassis for 472A1 apparatus case.

two 54-pair stubs in each case service the copper pairs in Coax 20. The size of the third stub varies depending on the cross-connect requirements peculiar to its station; these are 26 pairs for the basic, 101 pairs for the regulating, and 152 pairs for the equalizing repeaters.

Electro-magnetic protection has been provided with carbon block protectors on 50 pairs in the regulating case and all pairs in the equalizing case. In order to protect the critical circuits of the system for reliability, gas tube protectors have been provided on the ten pairs associated with the monitoring oscillator in all three apparatus cases.

In the shock testing program, the 472 type apparatus case survived shock loads of over 200 g.

4.3 *Cable Terminals*

4.3.1 *Coaxial Terminals*

The signal over the coaxials in the main cable is transmitted to the electronic equipment in the apparatus case by means of the 66-type cable terminals. Basically, these are gastight terminals using one, two, or four standard 0.375-inch disk-insulated air dielectric coaxials for use at the equalizing, regulating, and basic stations, respectively. The cable terminal for the basic stations is shown in Fig. 22.

The head of the terminal consists of a cylindrical tin-plated bronze housing with coaxial cable emanating from one end and a faceplate at the opposite end. A male screw connector is mounted on the faceplate for each coaxial. A circular trapezoidal groove for accommodating B sealing compound and four bolt holes in the faceplate

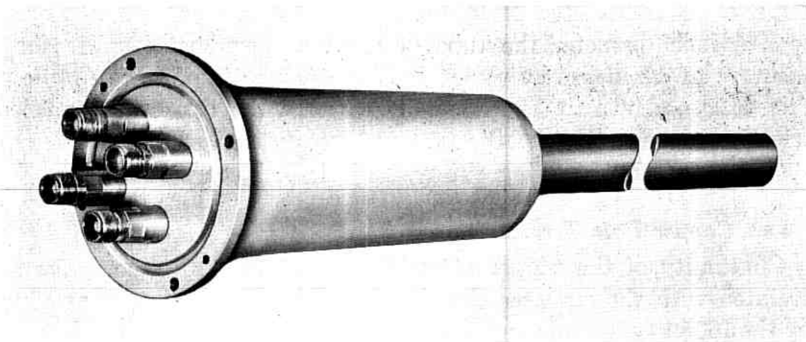

Fig. 22 — 66A1-4 cable terminal.

provide the means for attaching and sealing the cable terminal to repeater housings.

Internally, the inner conductor of the coaxial is spliced to an annealed copper wire which passes through the glass-bead seal located in the terminal head. The continuity of the outer conductor is maintained by means of a compensating sleeve which is crimped at the coaxial end and soldered to the male connector and the housing at the head end.

Electrical tests in the laboratory indicated that a 2 picofarad capacitive lump existed in the glass-bead area. The magnitude of the associated impedance discontinuity is shown in Fig. 23. A minor design change was subsequently made in the ⅝ inch splicing tube to

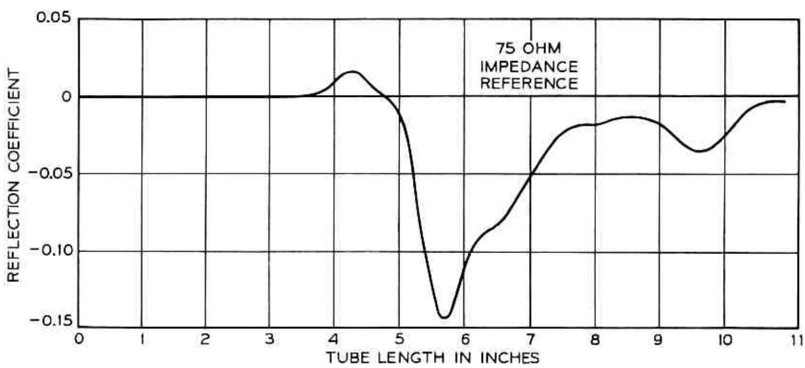

Fig. 23 — Impedance profile of the 66A1 type cable terminal with straight splicing sleeve.

improve its impedance uniformity. The modification was to expand to a $\frac{7}{8}$ inch diameter the tube at the inner conductor splice for a distance of $1\frac{1}{4}$ inches as shown in Fig. 24. The impedance profile of this design, shown in Fig. 25, indicates that the two areas are electrically compensating for the range of frequencies used in the L-type systems, thereby lowering the average impedance mismatch.

4.3.2 *Copper Pair Terminals*

Continuity of the copper pairs in Coax 20 to the copper pairs associated with the repeater chassis in the apparatus case, is provided by the 100A1 type cable terminals.

The 22 gauge PE-PVC insulated copper pairs in the cable are terminated on a female plug-in connector mounted on a tin-plated bronze faceplate. A gas dam is constructed in the Lepeth sheath cable stub to preclude the flow of air. The units attach to one of the ports in the base of the repeater case with four screws and sealed gastight at the interface with B sealing compound. A bypass valve facilitates control of gas pressure.

The 100B1-4 cable terminal, shown in Fig. 26, serves as the lineman's and cableman's order wire terminal and is mounted in the collar of the manhole for ready access from above ground. It is composed of a tin-plated bronze casting with a hinged cover and is sealed by means of an O-ring. A pressure plug is provided in the casting and the four 22 gauge PE-PVC insulated copper pairs in the Lepeth sheath cable stub terminate on two female plug-in connectors. Gas pressure is controlled by means of a bypass valve.

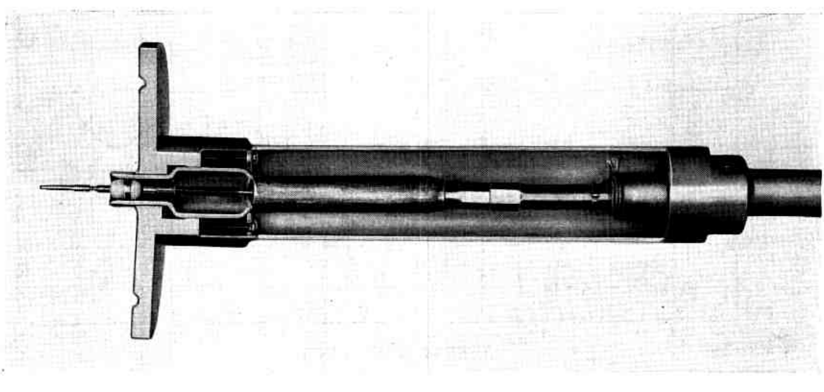

Fig. 24 — Cross section through head of 66A1-1 cable terminal with modified compensating sleeve.

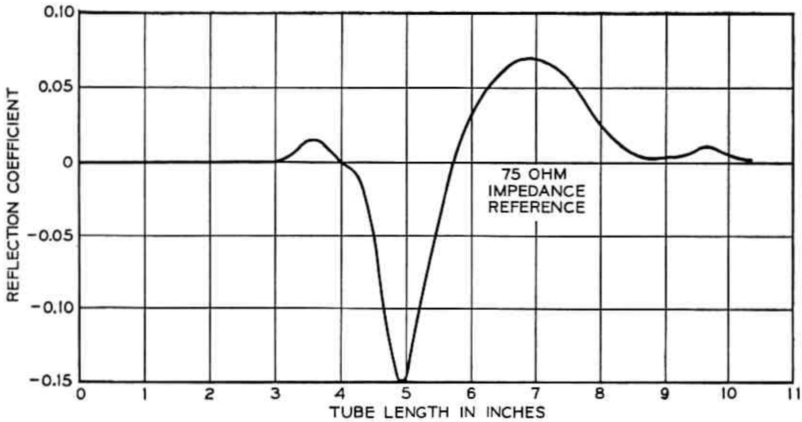

Fig. 25 — Impedance profile of 66A1 type cable terminal with modified compensating sleeve.

4.4 *Auxiliary Apparatus*

Although the main cable comprises 20 coaxials, economic considerations concerning initial plant investment required that apparatus be designed so that coaxials not initially in use could be equipped for service as warranted by system growth.

The repeater apparatus case, 471 type which represents a good size investment, has three ports for accepting coaxial and copper pair terminal facilities as discussed in Section 4.2.1.

To facilitate total racking of coaxial cables in the manholes and to provide a low cost item that could be substituted for the 471 type apparatus cases at repeater points, the 173A adapter, shown in Fig. 27 was made available. Basically, this is a galvanized cast iron "T" pipe having flanged openings that coincide dimensionally with the ports in the apparatus case. Continuity of the coaxial cable is maintained by patching through the adapter. When service is required, the terminating apparatus connections are broken, the adapter is removed, a 471 type apparatus case is installed on the bracket, and the connections remade.

V. RELIABILITY

Two fundamentally different approaches to achieving reliability are (*i*) to provide a very secure facility and (*ii*) to provide alternate standby facilities. Both approaches are being used. Economic considerations dictate that, since a full alternate facility would be prohibitively expensive, measures should always be taken to make the medium reliable. On the other hand, alternate standby facilities are provided at critical points along routes, such as river crossings. In the case of the L-4 system, reliability requirements have included hardening.

Many features are involved in hardening and, although hardening itself refers to the measures taken to increase chances of surviving a nuclear attack, most of the measures are effective in reducing the incidence of failures from ordinary causes. First, routes are chosen to circumvent major target areas and wires are placed above the cable in the trench to assist the sheath in carrying currents induced by an electromagnetic pulse as well as lightning. The cable is buried at greater depth than had generally been used in order to lessen shock and to reduce the hazards of accidental cable "dig-ups." And finally, there were requirements for ruggedness of components.

Also considered in designing for reliability were the slow wear out mechanisms of failure such as fatigue and corrosion. Although it is much too early to judge the effectiveness of design reliability, especially as regards wear out failures, early results indicate that the incidence of catastrophic failures, such as "dig-ups" by outside con-

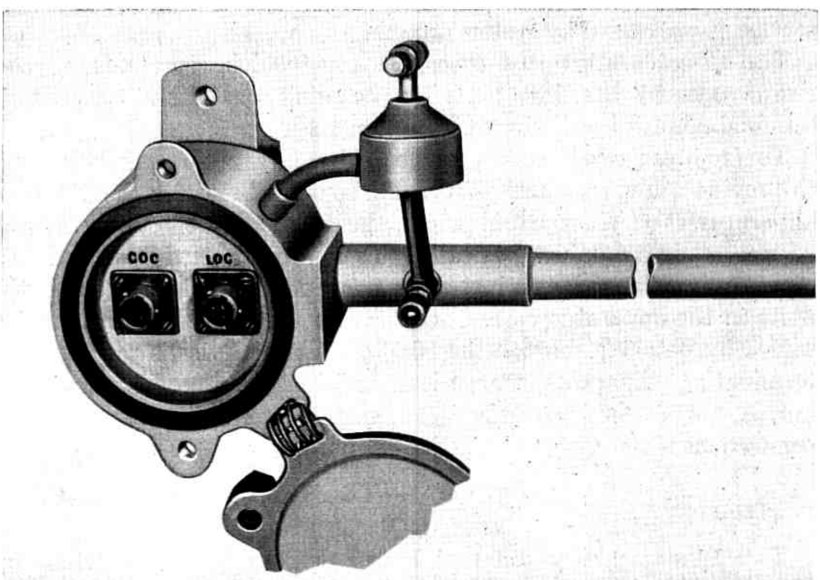

Fig. 26 — 100B1-4 cable terminal.

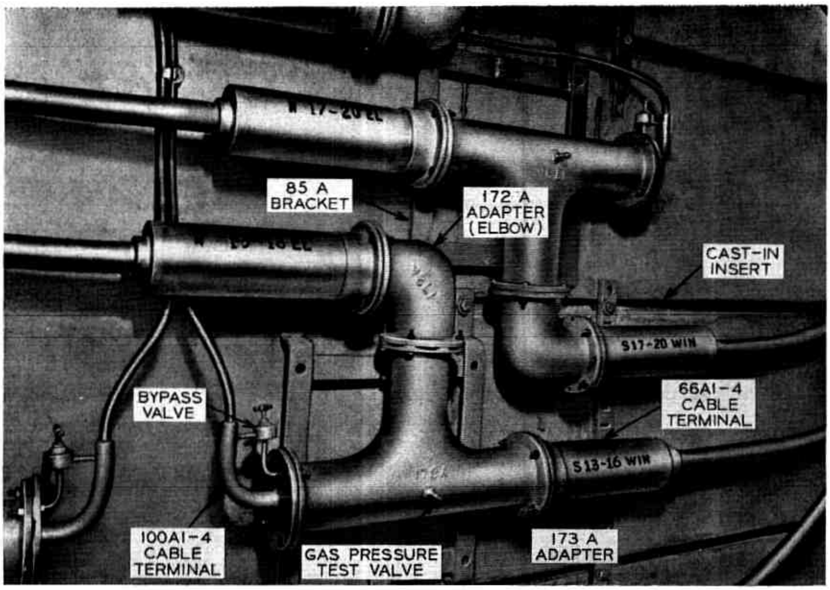

Fig. 27 — View of a typical manhole wall with dummy adapters.

tractors and lightning strokes, have been significantly reduced for the new hardened cable routes compared with similar but older facilities. At present, construction near roads and on farms appears to be the major cause of outages.

VI. SUMMARY

The overall success of the L-4 transmission system has stimulated much further work in the coaxial cable and apparatus design area. Needless to say, the features deemed responsible for the excellent performance of this system may be expected to be the basis for improved coaxial systems in the future.

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R. J. WIRTZ, B.S. (M.E.), 1950, Brown University; Bell Telephone Laboratories, 1956—. He has worked on resistor development, and design of N-3 carrier system equipment. He is supervisor of the L-4 Equipment Design Group, which is now working on physical designs for the L-5 coaxial carrier system.