THE BELL SYSTEM TECHNICAL JOURNAL

L5 Coaxial-Carrier Transmission System

Foreword

Coaxial-carrier transmission systems constitute a significant portion of the Bell System long-haul transmission facilities. These systems have been developed over several decades to provide the basis of a high-quality, high-capacity, long-distance communications network.

After extensive exploratory work on wideband amplifiers and coaxial cable, the feasibility of a coaxial-carrier system was demonstrated in 1936 between New York and Philadelphia. The success of this trial was followed by development of the first Bell System coaxial-carrier transmission system, the L1. Placed in service in 1941, the L1 system was initially capable of carrying 480 four-kHz two-way message channels per pair of 0.270-inch-diameter coaxial cables, with a repeater spacing of 5.5 miles. Soon, 0.375-inch-diameter cable became standard and, with system improvements, the vacuum-tube-operated L1 system was capable of carrying 600 circuits per coaxial pair with 8-mile repeater spacing. Its capacity was later increased to 720 circuits.

The major expense of the coaxial system has been in the outside plant area: cable, cable placement, right-of-way, and buildings, including aboveground or underground structures for housing repeaters. Once this portion of the system is established, development of electronic equipment to provide maximum utilization of the cable becomes economically attractive. Each successive generation of repeaters achieved wider transmission bandwidth and, hence, larger channel capacity through use of shorter repeater spacing, new technology, and more advanced system concepts. Use of cables with more

Table I — Evolution of coaxial-carrier transmission systems

System	First Service	Capacity in Telephone Circuits		Typical Coax Units Per Sheath	Approximate Repeater Spacing (Miles)	Repeater Technology	Repeater Types	Equalization	Other Key System Features
		Per Coax Pair	Per Cable*						
L1	1941	720†	720 2,160	4 8	8	Vacuum tube	Manually and automatically adjusted regulating repeaters; equalizing repeaters	Adjustable static and dynamic "bump" shapes	
L3	1953	1,860	5,580 9,300	8 12	4	Vacuum tube, statistically controlled key components	Buried-thermistor and line-pilot-controlled regulating repeaters; equalizing repeaters	Manual "cosine" shapes; dynamic broad shapes	Hardened configuration
L4	1967	3,600	32,400	20	2	Discrete transistor, printed wiring board	Fixed basic repeaters; regulating repeaters controlled by both a buried thermistor and line pilot; equalizing repeaters	Static "bump" shapes	Noise objective more stringent by 4 dB
L5	1974	10,800‡	108,000‡	22	1	Discrete transistor, hybrid integrated circuit	Same repeater hierarchy as L4	Static "bump" plus dynamic cause-related shapes	Phase-shaping networks to control third-order modulation addition; same noise objectives as in L4

* One coaxial pair reserved for protection of failed regular lines.
† Originally 480 circuits per coaxial pair, but widely used at 600-circuit capacity.
‡ Extensions to L5 are being developed to provide 13,200 telephone circuits per coaxial pair, or 132,000 circuits per 22-tube cable.

coaxial units per sheath increased route capacity and further reduced per-channel-mile costs.

From the outline of the evolution of coaxial-carrier systems in Table I, we can see that, in 33 years, the channel capacity of repeatered coaxial line has increased by a factor of 22.5—from 480 to 10,800 channels, and further increases are anticipated. In the same period, improvements in cable technology allowed the manufacture of cable with 5.5 times more coaxial units in the cable sheath—from 4 tubes to 22 tubes—resulting in a 10-fold increase in signal-carrying capacity, not including the two units reserved for service protection. The total impact, then, was a 225-fold increase in route capacity. During the same 33 years, Bell System circuit miles increased 485-fold.

This issue of *The Bell System Technical Journal* describes in detail the latest in the line of coaxial systems—the L5 Coaxial-Carrier Transmission System. The articles include descriptions of an advanced systems approach and sophisticated repeatered-line and equalization designs. Also included are the novel concepts in repeatered-line powering, line-protection switching, equipment-performance surveillance, centralized maintenance, and carrier reference-frequency generation. Other articles describe the new multiplex and signal-administration equipment and the important role in the success of L5 played by innovations in physical design and thin-film techniques and by the use of ultralinear semiconductor devices. The many computational aids and measurement facilities that were effectively used in the development of the system are also discussed.

The initial L5 system—815 miles of cable, 14 stations, 850 manholes, 3400 manhole repeaters, and over 250 bays of transmission equipment—went into service on January 3, 1974, fulfilling a schedule developed six years earlier. This on-time completion of such a massive system required the dedicated effort of many individuals in the Bell System companies. Bell Laboratories people conceived and developed the system and its components; Western Electric people were responsible for manufacture of cable and electronic equipment and for installation of main-station equipment; and AT&T and Long Lines people were actively involved in system planning, coordination, route selection, cable placement, installation of line equipment, and operational testing. It is to this skilled Bell System team that this issue is dedicated.

L5 System:

Overall Description and System Design

By F. C. KELCOURSE and F. J. HERR

The L5 Coaxial-Carrier Transmission System is a new solid-state broadband system designed to transmit 10,800 long-haul message channels on standard $\frac{3}{8}$-inch (9.5-mm) coaxial cable. Alternatively, high-speed digital signals may be transmitted in selected mastergroups. This paper contains an overall system description and describes in detail system design considerations such as the transmission plan, frequency allocation, and 4000-mile noise performance. Also considered are the significant properties of the coaxial medium, the ways they affect the hierarchy of equalizers traditionally found in long-haul coaxial systems, and the manner in which the several levels of equalization relate to each other and to the fundamental transmission phenomena within the system.

I. INTRODUCTION

During the last 30 years, long-haul transmission facilities have been rapidly expanded to meet the increasing need for both voice and data transmission, and it is expected that rapid growth will continue in the 1970's although possibly at a slower rate than the 15 percent per year which typified the late 1960's. Thus, during this period there has continued to be a need for new systems providing increased message capacity at lower cost to supplement the systems already available in the long-haul plant. As can be seen in Table I, the L5 Coaxial-Carrier Transmission System is the latest in a family of coaxial transmission systems developed to date.*

It is evident that the newer coaxial systems have achieved increased capacity because of better utilization of the transmission medium (i.e.,

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* The sheath capacities listed in Table I are the ones most commonly associated with each of the systems. The initial L3 application, in 1953, was actually on Coax-8 (i.e., a cable which includes eight coaxials) with a total sheath capacity of 5580 circuits on three working pairs, with one pair reserved for protection.

Table I — Some features of L-type systems

	L1	L3	L4	L5
First service	1941	1953	1967	1974
Technology	Tubes	Tubes	Transistors	Transistors and HIC's
Repeater spacing (miles)	8	4	2	1
Channels per pair	600	1,860	3,600	10,800
Total channels/sheath	1,800	9,300	32,400	108,000
Noise objective (dBrnc0)	44	44	40	40

wider bandwidth) and because of an increase in the number of coaxials incorporated within the sheath. An important characteristic of such long-haul systems is the dominant effect which the installed cost of the transmission medium has on the overall cost. By providing triple the message capacity on L5 as compared to L4, for example, the cost of a message-circuit mile on L5 is substantially less than that of a message-circuit mile on L4, even though the number of repeater stations is doubled and the electronics required is of wider bandwidth and of increased sophistication. In addition, the older L-carrier systems use the same transmission medium, and it is possible to convert many existing L1, L3, or L4 routes to L5 by adding the required number of new repeater sites to the existing sites of the earlier system. In this manner, increased message capacity can be obtained at an incremental cost that is even more attractive than on a new installation.

As noted in Table I, the L5 coaxial system provides 10,800 toll-grade quality message channels (on a pair of 0.375-inch disc-insulated coaxial cables). In a Coax-22 cable configuration, the resultant total channel capacity is 108,000 channels per cable sheath when a 1:10 protection strategy is employed. The increase in the bandwidth of the system required for this capacity is made possible in part by setting the nominal repeater spacing at 1 mile and in part by several significant technological advances in such areas as repeater, network, and device design.[1-3]

The L5 system includes fixed gain "basic" repeaters at 1-mile nominal intervals, adaptive "regulating" repeaters at 7-mile maximum intervals, adjustable gain "equalizing" repeaters at 38-mile maximum intervals, and power-feed main station repeaters at 75-mile maximum intervals.[4] Line protection switching is provided on a 1-for-10 basis at maximum intervals of 150 miles.[5]

The adjustable equalization of the L5 repeatered line is partially manual and partially automatic. Manually adjusted equalizers compensate for the transmission deviations, which are essentially time-invariant. These equalizers are adjusted to minimize the mean-squared error of the broadband channel upon initial installation, and any subsequent readjustments are normally done out of service. The use of out-of-service, mean-squared-error adjustment procedures permits a substantial improvement in the transmission quality for a particular array of adjustable equalizers.

Automatic equalizers compensate for the time-variant transmission deviations. These equalizers are located in each regulating, equalizing, and main station repeater and fall into two categories. The automatic equalizers for the cable-temperature effect are located in each of these repeaters and utilize both open-loop regulating circuits, which detect the resistance of a locally buried thermistor, and closed-loop regulating circuits, which detect the level of a full-time line pilot. The automatic equalizers for the repeater-temperature effect and the regulating repeater tracking error are located only in the receiving main stations and utilize closed-loop regulating circuits equipped with "memory," which detect the levels of four full-time line pilots.

Provisions are included at appropriate main stations for adding, dropping, or branching any jumbogroup (3600 channels) at line frequency.[6] "Dropping" involves connection to the jumbogroup multiplex (JMX), a new frequency-division multiplex (FDM) terminal designed for L5, which translates up to three basic jumbogroups (0.564 to 17.548 MHz) to appropriate line frequency assignments between 3.124 and 60.556 MHz.[7] "Branching" involves connection, at line frequency, to or from another L5 facility. "Adding" involves the addition to an L5 line of circuits originating from the JMX. The mastergroups of a basic jumbogroup are administered much like the mastergroups of an L4 signal. The basic jumbogroup can consist of any available combination of analog and digital mastergroup signals. When frequency assignments permit, direct connection with other facilities is made possible by connecting all mastergroups to the JMX via a basic jumbogroup trunk bay (BJGT). This trunk bay permits mastergroups from several different sources, such as mastergroup multiplex (MMX-2), L-carrier mastergroup digital (LMD) terminals, and radio systems, to be combined into a basic jumbogroup.[6] The carrier generation within the JMX requires reference frequencies of an accuracy heretofore unavailable in the Bell System and consequently

a jumbogroup frequency supply (JFS) is included at all JMX locations for this purpose.[8]

At selected main stations, an L5 transmission surveillance center (TSC) will be utilized to centralize and partially automate the surveillance and fault location of several hundred miles of repeatered line and associated main station and multiplex equipment.[9] In the initial L5 installation between Lillyville, Pa., and Hillsboro, Mo., a single TSC serves these functions for 664 miles of "backbone" route Coax-22, and 151 miles of "sideleg" Coax-12.

II. GENERAL FEATURES OF THE L5 SYSTEM

All long-haul coaxial systems share the common objective of toll-grade signal-to-noise quality in a 4000-mile circuit. In the L5 system, this involves equalization of over 100,000 dB of cable loss at the uppermost message frequencies and almost 132,000 dB at the uppermost "controlled" frequencies near 70 MHz. Equalization of this loss is accomplished with a combination of wideband circuits, all of which must be low in noise, highly linear, and extremely stable. It is in the achievement of the 4000-mile transmission objective that the familiar coaxial system hierarchy of increasingly complex repeaters evolves. Among the general features of the system, the frequency allocation will be considered first. Although many different factors must usually be considered during the specification of a frequency allocation, only a few can be reviewed here.

2.1 Frequency allocation

The controlling consideration in establishing the frequency allocation is the overall bandwidth involved—both in terms of hertz of bandwidth and in terms of the ratio of uppermost to lowermost frequencies to be used—which appears to be consistent with the state of the art at the time, or within the evolutionary lifetime of the system development. An important and relevant experience of previous coaxial-systems development has shown the difficulty and expense of simultaneously controlling the transmission characteristic at the highest and lowest frequencies of a spectrum that spans a frequency range of as much as five to six octaves. Realizable transformers, inductors, and capacitors are in most cases far from ideal components, and parasitic effects become especially troublesome when the frequency range is more than a few octaves. As the system bandwidth approaches five octaves or so, parasitic effects frequently become dominant. Thus, for a given band-

width, there may be significant advantages to raising, within reasonable limits, the lowest frequency requiring careful control.

Compatibility with the existing FDM hierarchy is another relevant factor and, in the case at hand, the basic building block was taken to be the six-mastergroup assembly formed in accordance with the L4 transmission plan. The six mastergroups are arranged between 0.564 and 17.548 MHz and are designated the "basic jumbogroup."

A basic system objective was the transmission of at least three times the L4 message capacity of a single jumbogroup. The assignment of the jumbogroup line frequencies and the spacing between the jumbogroups depends mainly on the bandwidth considerations already discussed and the types of signals and signal administration procedures planned for the system. In L5, the capability to select any jumbogroup from the line spectrum *at line frequency* ("branching and dropping") was a basic system objective. Questions relating to the design of appropriate bandpass and band-stop filters substantially affect the guard spacing required between jumbogroups. In addition, at the time that it was necessary to make the frequency allocation decisions, the ability to displace an entire analog jumbogroup with a high-speed digital signal was also a basic objective.

The frequency allocation of the L5 system is shown in Fig. 1, with six mastergroups, each containing 600 message channels, making up the basic jumbogroup. In some cases, a mastergroup in the L5 line spectrum will transmit a 13.29-megabits-per-second digital signal by means of the L-carrier mastergroup digital (LMD) terminal,[10] a digital repeater spaced at maximum intervals of 300 miles. These digital channels will be used to carry a variety of digital signals including wideband data, DDS, and digitized *Picturephone*® signals.

The regions around 1.59 and 68.65 MHz are used to transmit fault-locating signals which in turn are used to remotely locate repeater faults from a centralized location equipped with an L5 transmission surveillance center.

The region around 68.78 MHz is used to transmit signals required by the line-protection-switching system (LPSS-3) for interstation communication, and control and verification of LPSS operation.

Pilots are transmitted below, between, and above the jumbogroups to control the L5 automatic equalizers. The automatic equalizers for the cable-temperature effect, located in the regulating repeaters, are controlled by the incoming level of the 42.880-MHz pilot (which is often referred to as the "temperature pilot"). The automatic equalizer
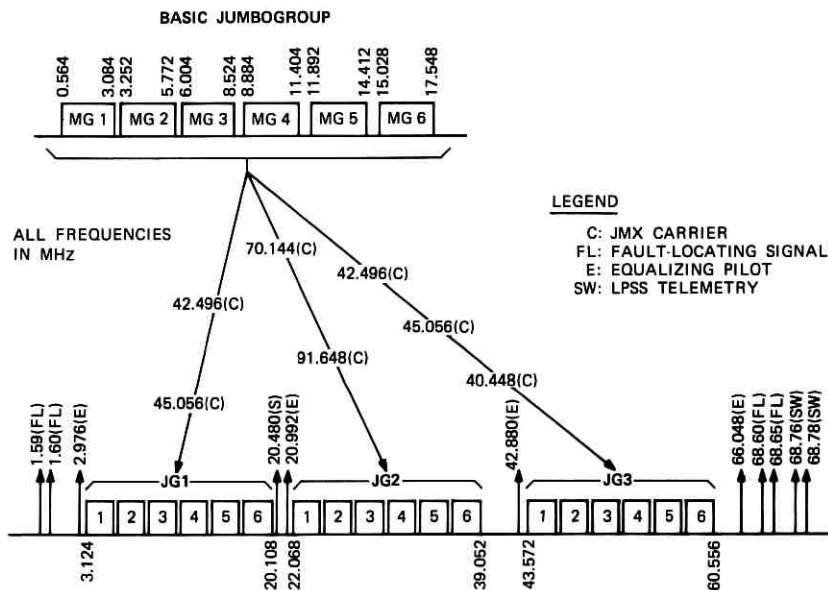
Fig. 1—L5 frequency allocation. The basic jumbogroup shown is equivalent to the L4 line spectrum. The carrier frequencies indicated are those used to translate the basic jumbogroup to the L5 line frequency assignment.

for the repeater-temperature effect and the regulating repeater tracking error, located in receiving main stations and designated the E3 equalizer, is controlled by the incoming levels of all four pilots.

After the establishment of at least an approximate frequency allocation, the central concept in the design of the coaxial system becomes the equalization of the 4000-mile cable loss. Furthermore, it must be done in such a way that the residual transmission deviations are less than 0.01 percent of the top channel cable loss, while accumulated noise from all sources, including associated multiplex, must satisfy the long-haul noise objective of 40 dBrnc0. The general features of the system layout selected to meet these goals are described next.

2.2 System layout

The most general features of a repeatered-line layout in long-haul-coaxial systems depend mainly on power-feeding techniques, equalization strategy, reliability requirements, and convertibility of older systems. The equipment providing power for the remote repeaters, the fine-grained equalization of transmission, and the automatic line-protection

switching will generally be of such size and sophistication as to require a station environment as opposed to the manholes that are suitable for the simpler repeaters. The cost of these main stations is many times that of a manhole, and it is essential that the number of such stations be minimized for system costs to be competitive with alternative ways of providing the circuits.

The power-feed configuration maximizing the length of the power span for a given voltage is the double-ended connection shown in Fig. 2.

The L5 line power supplies provide a nominal 0.91 A dc via the center conductors of the coaxial cables for powering up to 75 miles of remote line repeaters. This is a convenient submultiple of the L4 power span length and requires a maximum line voltage of only ±1150 V, with the result that physical designs of high reliability should be readily achievable, and corona "popping" in the installed system should be minimal.

The basic layout is affected by the equalization strategy mainly by the effect the latter has on repeater signal loading and the extent of signal level misalignment that must be accepted by the line repeaters. The L5 strategy in most cases uses pre- and postequalization in equal parts and, as a result, repeater load requirements can be reduced for a given transmission deviation from nominal or, for a fixed repeater load capacity, the permissible distance between equalizers can be increased, since the magnitude of the transmission deviation is usually
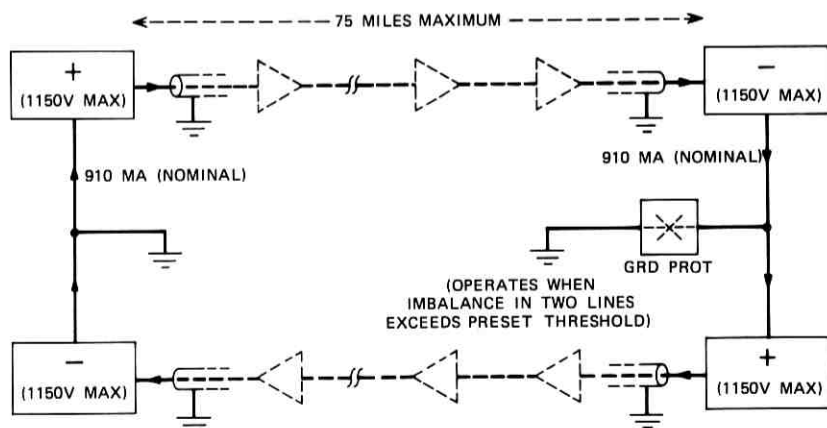

Fig. 2—Simplified diagram of line power feed. Plus or minus voltage is developed by dc-dc converters and combined with message in power separation filters prior to application to coaxial line.

proportional to distance. Finally, when the misalignment is contained to within about ±5 dB, it can be shown that this approach achieves optimum signal-to-noise ratio in both second- and third-order inter-modulation-limited systems.[11] Figure 3 shows the signal-to-thermal noise penalties associated with single-ended equalization, at either the receiver or the transmitter only, and with equal parts of pre- and postequalization.

Finally, a relatively simple adjustable equalizer is included in the midspan manhole repeater so that considerations of equalization and load capacity do not entirely determine the main station interval. The midspan equalizer provides coarse adjustment of signal levels as they traverse the power feed span, and not the fine-grain compensation of transmission deviations, which is accomplished only at the main stations.

Pre- and postequalization is applied throughout the L5 repeatered line except for the automatic equalization of the repeater-temperature effect, the controlling considerations in this case being simplification of system operation and reduction of system complexity. The repeater-temperature effect is equalized only in the receiving portion of main station repeaters, which are up to 75 miles apart.

Reliability affects system layout, since the reliability of the system electronics determines the length of the protection-switching span and
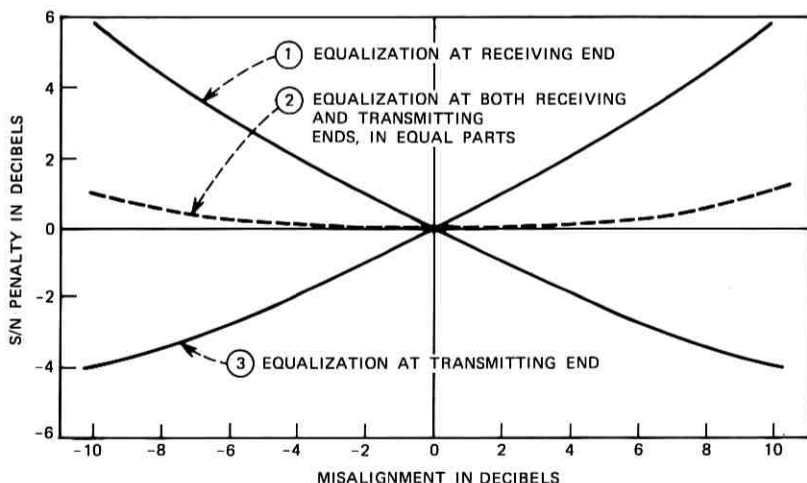

Fig. 3—Signal-to-thermal noise penalties resulting from signal misalignment. For misalignment less than about ±5 dB, pre- and postequalization in equal parts (curve 2) also results in negligible penalties in second- and third-order intermodulation noise (not shown).
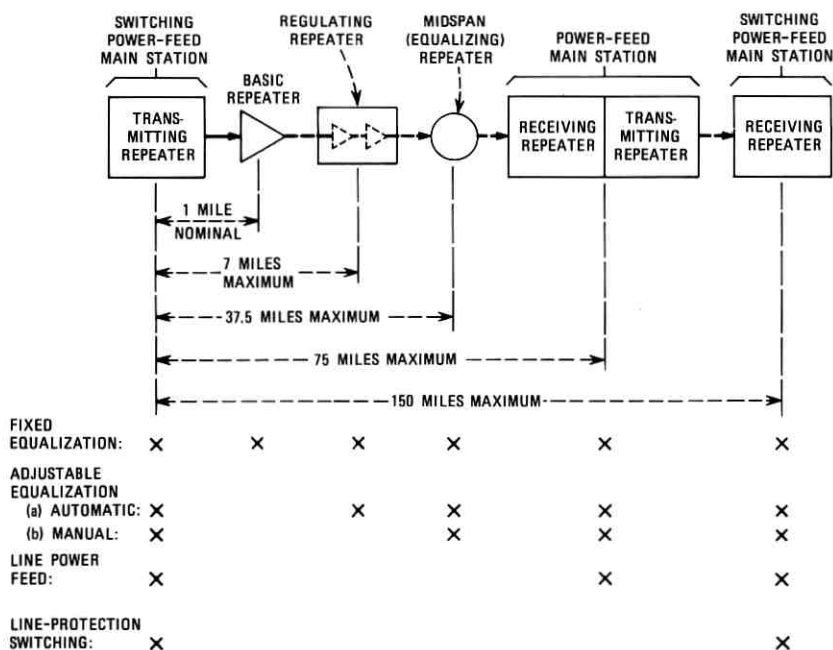
SWITCHING POWER-FEED MAIN STATION — TRANSMITTING REPEATER

BASIC REPEATER

REGULATING REPEATER

MIDSPAN (EQUALIZING) REPEATER

POWER-FEED MAIN STATION — RECEIVING REPEATER / TRANSMITTING REPEATER

SWITCHING POWER-FEED MAIN STATION — RECEIVING REPEATER

1 MILE NOMINAL

7 MILES MAXIMUM

37.5 MILES MAXIMUM

75 MILES MAXIMUM

150 MILES MAXIMUM

	Transmitting repeater	Basic repeater	Regulating repeater	Midspan repeater	Power-feed main station	Switching power-feed main station (receiving)
FIXED EQUALIZATION:	×	×	×	×	×	×
ADJUSTABLE EQUALIZATION (a) AUTOMATIC:	×		×	×	×	×
(b) MANUAL:	×			×	×	×
LINE POWER FEED:	×				×	×
LINE-PROTECTION SWITCHING:	×					×

Fig. 4—Main features of an L5 switching span.

the number of lines required as standby for any given availability objective to be achieved. The predicted reliability of the L5 line repeaters permits up to two power-feed spans to be traversed before automatic line-protection switching is required, and then on a 1-for-10 basis. In the context of the repeatered line, the basic building block of the system is the switching span and, in the absence of signal administration needs (dropping, branching, etc.), such spans as are shown in Fig. 4 are connected in tandem until the distance between terminal stations is bridged.

In addition to the provision of repeatered-line power, fine-grain equalization, and (sometimes) line-protection switching, other important functions are carried out at the different types of L5 main stations, and, depending on the particular features required of main stations, they fall into one of three categories. In order of decreasing complexity they are:

(i) Terminal station/terminal main station (TS/TMS).
(ii) Switching power feed main station (SPFMS).
(iii) Power feed main station (PFMS).

The functions performed at these stations which are directly related to the repeatered line are alike. The differences among them are in the signal and pilot administration (line connecting) functions, the line-protection switching, the transmission-surveillance capabilities, and the multiplexing and related signal-processing functions.[6] Some main station and repeater functions are summarized in Fig. 4, and some possible terminal interconnections for jumbogroup administration are shown in Fig. 5.

2.3 System installation, operation, and maintenance

A basic objective in the design of the L5 system was to achieve a system requiring a minimum of day-to-day routine involvement of craftspeople. The L5 main stations, as well as the repeatered line, will generally be operated without personnel present. This has been made possible in part by developing equalization procedures that should require activity only rarely, and in part through the implementation of centralized alarm, surveillance, and fault-locating techniques. The layout of the initial L5 commercial installation between Lillyville, Pa., and Hillsboro, Mo., is shown in Fig. 6 and, for purposes of the following discussion of system organization and administration, is treated as typical.

2.3.1 Installation and maintenance

Installation of the L5 repeatered line is accomplished by telephone company personnel and is controlled and supervised at "maintenance centers." Six such maintenance centers are shown in Fig. 6; the four at Danville, Xenia, Morgantown, and Noble have responsibility for all but the 30 repeater sections near Lillyville and the 14 repeater sections near Hillsboro. Those four stations are consequently each responsible, on the average, for the installation and maintenance of nearly 200 miles of repeatered line which, when fully equipped, will include approximately 4000 line repeaters.

Upon being received from the factory, every L5 repeater is carefully tested at one maintenance center before installation along the repeatered line. The tests made involve primarily transmission measurements that are executed, on site, by a computer-operated transmission measuring set and associated test stand. A hard copy of the test results is produced during the process and, in the case of an acceptable repeater, is retained in the maintenance center as part of the permanent history file on that repeater. Regulating repeaters and fault-locating oscillators are among system components requiring on-site adjustment
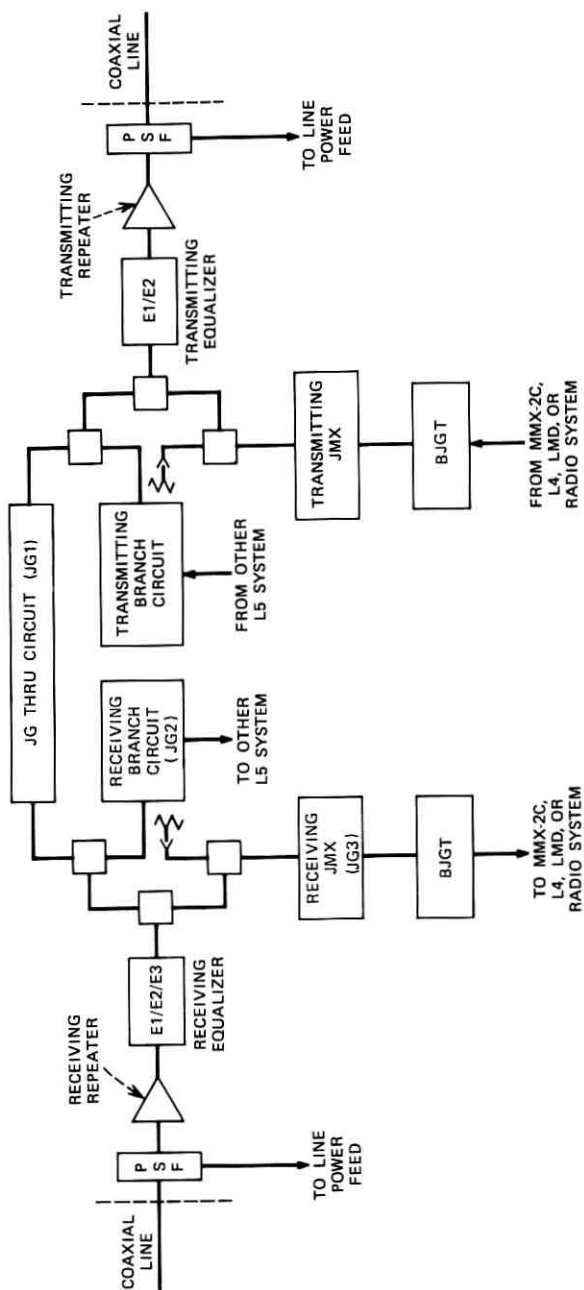
Fig. 5—Jumbogroup administration. In the hypothetical case shown, JG1 is through-connected at line frequency and JG2 is branched on to another L5 system also at line frequency. JG3 is dropped via JMX and the BJGT to basic jumbogroup for interconnection with MMX-2, L4, a radio system, or a combination thereof.
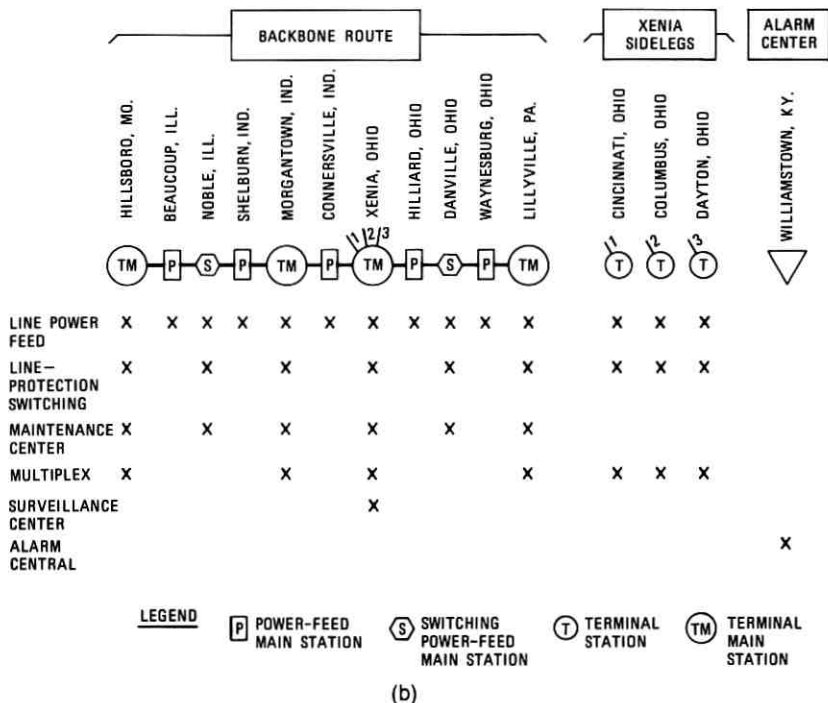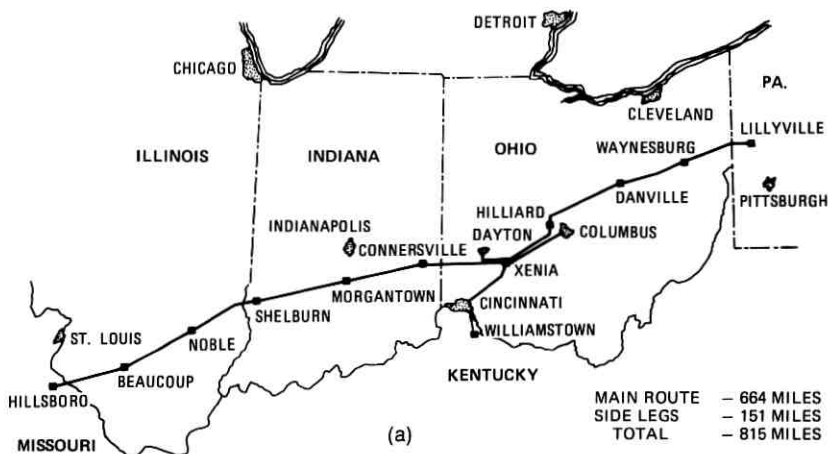
Fig. 6—(a) Geographical layout of initial L5 commercial installation between Lillyville, Pa., and Hillsboro, Mo. (b) Administrative layout of Lillyville-Hillsboro route.

in accordance with the particular repeater station to which they are assigned, and those adjustments are normally made utilizing the same test setups.

Installation of the L5 main station equipment is usually done by Western Electric installation personnel and, prior to turnover of the equipment to the telephone company, a variety of transmission, noise, and operational tests is usually necessary. In most cases, these tests utilize the same kind of test equipment required by the telephone company to install, maintain, and operate the system.

After the testing and installation of the repeaters and fault-locating oscillators have been completed between a pair of power-feeding points (such as Lillyville and Waynesburg in Fig. 6), and after the main station equipment has been installed and tested, power is applied by operating the line-power-feed dc-dc converters at both ends of the line. The converters at each end are adjusted to share equally the load of the intervening repeatered line (except in the case of very short power spans, which can be powered by converters at one end only). With power applied to the line, a detailed step-by-step process, known as the L5 line acceptance procedure and appropriately documented, is executed, which permits the systematic identification and elimination of any system faults that may have escaped detection during the earlier testing of the individual repeaters and cable sections. In addition, such engineering or operational errors as incorrect line-build-out specification, incorrect fault-locating oscillator adjustment, or improper pre-regulator settings can, with a combination of engineering and on-site analysis, readily be detected. The line acceptance procedure flows from a detection of the gross faults that may have existed, through a regulator-by-regulator pilot level measurement sequence, to the measurement, evaluation, and equalization of the channel between main stations, and ultimately to the noise loading of the tandem power-feed spans making up a protection-switching span (such as between Lillyville and Danville in Fig. 6). Upon completion of this procedure, the repeatered line is ready for the next level of system testing, which would consist of such procedures as exercise and validation of the protection-switching system, including tandem operation (such as from Lillyville-to-Danville-to-Xenia in Fig. 6) and multiplex-to-multiplex measurements and evaluation (such as between Lillyville and Xenia).

2.3.2 System alarms

A wide variety of status and alarm indications has been made available in the L5 system design to provide for complete and ready

remote verification of system health as well as diagnosis of system troubles when they occur. Status indications are provided for such operating parameters as the dynamic equalizer range setting and the condition of the protection-switching system, while alarm indications are developed based on the power level of any line or JMX pilot, the condition of any fuse, and a variety of other parameters. These alarm and status indications are transferred to a central alarm station which, in Fig. 6, is located at Williamstown, Ky., off the L5 route. This station is part of the E2 Status Reporting and Control System and, in the following discussion, references to E2 will be to this system, not to the E2 equalizer that is an integral part of the L5 system. Transmission between the L5 stations and the E2 central station takes place over a data network which is a four-wire private line, meeting requirements similar to those of a standard voice-band data service. (In the case of the initial L5 installation, the data facility is located internal to the co-axial cable, utilizing two of the loaded 19-gauge pairs, as far as Cincinnati, and utilizes an external private line circuit between Cincinnati and Williamstown.) Each L5 main station is equipped with at least one E2 "remote unit," which provides the interface between the L5 system alarms and the E2 data facility. The larger L5 stations, such as Xenia and Morgantown on the initial route, require two E2 remote units, because of the quantity of equipment to be located there. The total load on the Williamstown E2 central is 16 remote units, the upper limit for a "manual" E2 central on a single data network.

In its normal operating mode, the E2 central sequentially interrogates (alarm polls) each remote unit for its alarm poll response. When all remote units have been polled, the process is successively repeated. A manual operation at the central or a demand by the L5 transmission surveillance system (TSS) for use of the data facility will interrupt this otherwise automatic and repetitive alarm polling. The manual operations at the central include calling up a particular station display for a detailed alarm report as well as controlling such L5 system operations as effecting a line switch or operating the line-feed converters.

2.3.3 Surveillance and fault location

The surveillance and fault-location features of the system are described in detail in Ref. 9, and will be touched on here only briefly. The addition of a remote call-up feature to the E2 Status and Alarm Reporting System described above permits the L5 TSS to time-share the E2 data facility and the multiple-address capability of the E2 central. The TSS of the initial L5 route has a single transmission-

surveillance center (TSC), located at Xenia. Equipped with a mini-computer central processor, and loaded with appropriate data-base, operating-system, and applications programs, that TSC controls the fault-location and surveillance processes for all the repeaters and main stations in the initial system. (The total mileage represented by Fig. 6 is approximately 815 miles. When fully equipped, the equipment "domain" of the TSC at Xenia will include over 17,000 line repeaters and more than 250 transmit-receive bays, as well as associated multiplex.) Virtually all the periodic maintenance measurements planned for the system will be carried out at the TSC and, by supplying a full duplex data circuit between Xenia and Williamstown, operation of the TSC can be transferred to Williamstown during Xenia's unmanned intervals.

III. DESIGN OBJECTIVES AND SIGNAL-TO-NOISE ANALYSIS

3.1 Repeater design objectives

The transmission plan of an analog coaxial system is determined by adjusting the bandwidth, repeater spacing, and signal transmission levels so that prescribed channel objectives are realized over the life of the system. To transmit over a particular bandwidth on a specified cable, the repeater spacing is determined by the available gain, noise figure, load-carrying capacity and linearity of the repeater, and the margins required to allow for cable and repeater variations resulting from aging, temperature changes, and manufacturing limits.

Table II illustrates how the basic repeater noise, load capacity, and linearity design objectives became progressively more stringent as coaxial-system design proceeded from L3 through L5. In comparing

Table II — Design objectives for L-type repeaters

System Data	L3	L4	L5
Top frequency (MHz)	8.3	17.5	60.6
Repeater performance, referred to top channel:			
Noise figure (dB)	11	6.5	5.5
Load capacity (dBm)	16	21	24
Nonlinearity			
Second order (dB)	−61	−70	−70
Third order (dB)	−96	−100	−110*
Insertion gain (dB)	44	33	31
Transmission level (dB)	−11	−14.2	−13.6

* This requirement applies only with the use of the phase-shaping networks described in Section 3.2.

the repeater design parameters of the table, it is important to note that the required noise, load capacity, and linearity improvements from L3 through L5 are achieved at progressively higher top channel frequencies. Table II shows, for example, that operating at one-half the L4 repeater spacing and more than three times the L4 top channel frequency, the L5 basic repeater requires a 1-dB better noise figure, 3-dB more load capacity, and 10-dB better third-order nonlinearity than the L4 basic repeater, with all parameters referred to their respective top channel frequencies. Top channel insertion gain is slightly lower in the L5 repeater despite the higher top transmitted frequency because of the 2:1 reduction in repeater spacing.

3.2 Signal-to-noise analysis

The L5 system signal-to-noise design objective is based on meeting a 4000-mile per channel noise objective of 40 dBrnc0. The 40-dBrnc0 total is allocated as 39.4 dBrnc0 to the high-frequency line and 31.5 dBrnc0 to the terminals (Fig. 7 shows how the 31.5 dBrnc0 terminal allowance relates to a possible multiplex deployment). In the signal-to-noise design of the system, high-frequency line transmission levels are shaped with frequency to realize noise performance in every channel as nearly equal as possible. In L5, second-order modulation
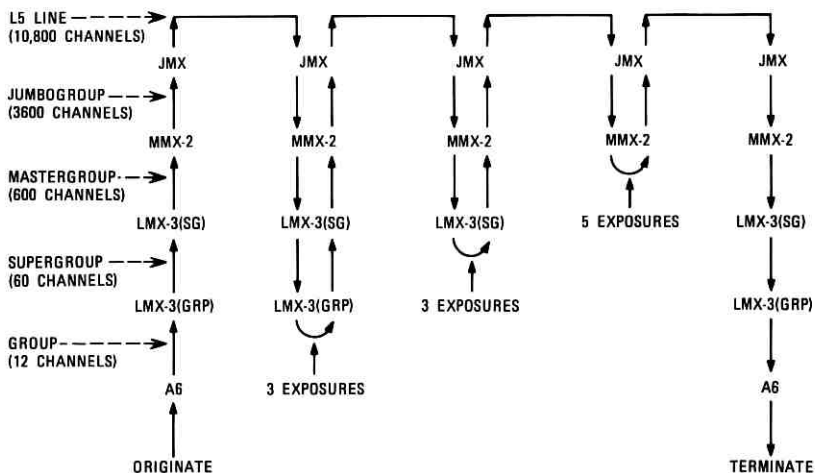

Fig. 7—Possible multiplex terminal deployment in a long-haul interconnection. This example includes 11 intermediate multiplex operations between the originating and terminating stations, which is well above the number typically encountered.

noise is the controlling modulation contributor because of the use of the phase-shaping networks discussed below, which serve to reduce the third-order intermodulation distortion to a relatively insignificant level. In L5, application of signal level shaping techniques results in the transmission levels shown in Fig. 8a and in the predicted zero level noise shown in Fig. 8b. The curve labeled "Total" in Fig. 8b is the predicted noise under nominal conditions. The uppermost curve, labeled "Total (with allowances)," is the predicted noise performance including penalties for misalignment, misequalization, and the like. As can be seen, the average noise over all channels is predicted to be 39.7 dBrnc0 under nominal conditions. The use of frequency frogging of the jumbogroups and mastergroups will tend to cause the typical 4000-mile L5 channel to approach that average value in noise performance. In addition, measurements on the L5 field trial and first commercial installations indicate close agreement with 4000-mile predicted noise proportioned to the appropriate test system length. A controlling factor in the determination of the allowable second- and third-order modulation coefficients listed in Table II is the use of repeatered line phase-shaping networks in each L5 regulating repeater. Such networks are used to produce a significantly less linear repeatered line phase-frequency characteristic, resulting in reduced correlation of third-order $(A + B − C)$-type product addition from repeater to repeater. In the absence of phase-shaping, such product addition would occur on a nearly in-phase or voltage-addition basis, with the repeater-to-repeater terms approaching $20 \log N$, where N is the number of tandem line repeaters. *

In the presence of the phase-shaping networks, such as those used in L5, substantially lower correlation of $(A + B − C)$ product addition results. For the L5 6-mile regulating-section phase-frequency characteristic shown in the upper curve of Fig. 9, the repeater-to-repeater addition term within a frogging link is approximately $15 \log N$, corresponding to a per-frogging-link $(A + B − C)$-type product reduction of about 15 dB relative to the "no-phase-shaping" values. Thus, in L5, the use of phase shaping reduces third-order intermodulation noise to relatively insignificant values, and the system is second-order modulation and thermal-noise limited.

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* Product addition according to a particular law occurs over "frogging" link intervals of up to 800 miles. Because of the randomizing effect on product and source frequencies produced by frogging, random or power addition is assumed to occur from frogging link to frogging link for all modulation product types.
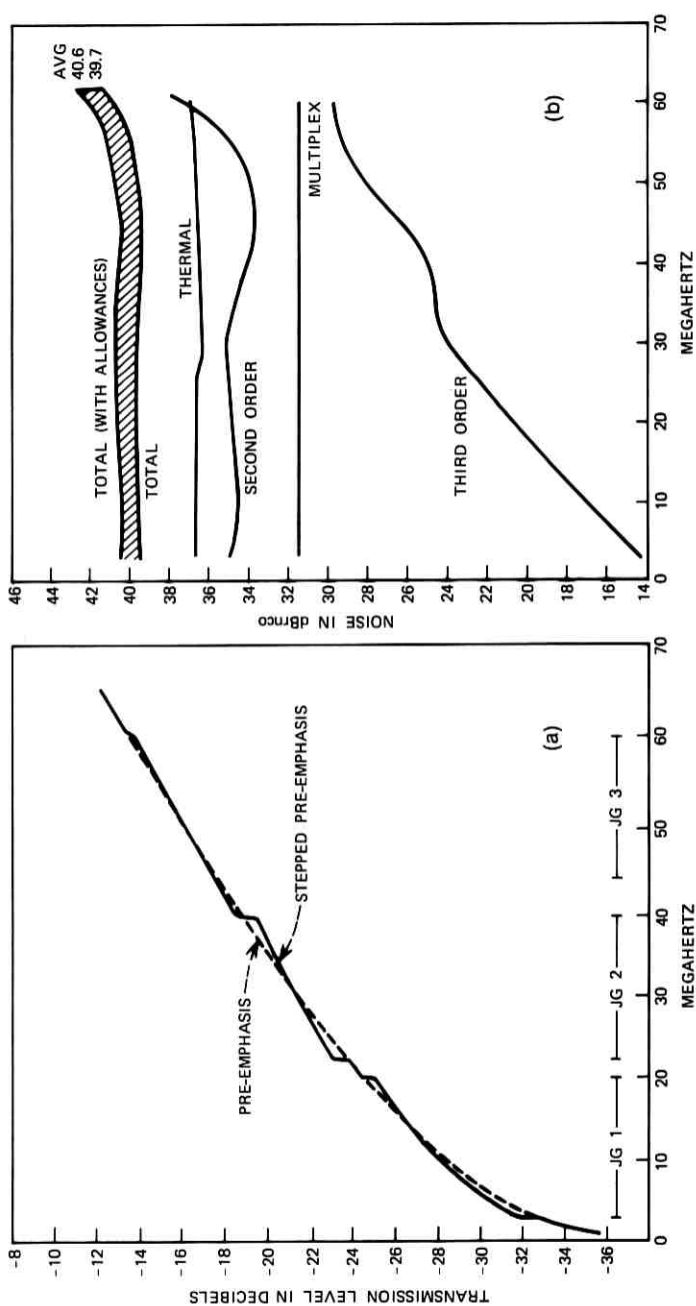
Fig. 8—(a) Repeater output transmission levels. The discontinuities in the TL shown result from using a stepped approximation to the optimum (smooth) preemphasis curve, derived by transmitting JG2 1.7 dB hotter than JG1 and JG3 1.8 dB hotter than JG2 at the JMX output. (b) Predicted zero level noise for 4000-mile L5 system. The prediction is based on extensive individual repeater and transmit-receive bay measurements, and has been confirmed by noise loading measurements made on lengths of the initial system up to 750 miles long.

PHASE SHAPING NETWORK (SIMPLIFIED)

$$L_1 \; \alpha \; \frac{1}{f_c b}, \quad C_1 \; \alpha \; \frac{b}{f_c},$$

$$L_2 \; \alpha \; \frac{b}{f_c}, \quad C_2 \; \alpha \; \frac{b}{f_c(b^2-4)},$$

FOR THE CURVE SHOWN
$b = 6$, $f_c = 65$ MHz

WITH PHASE SHAPING

WITHOUT PHASE SHAPING
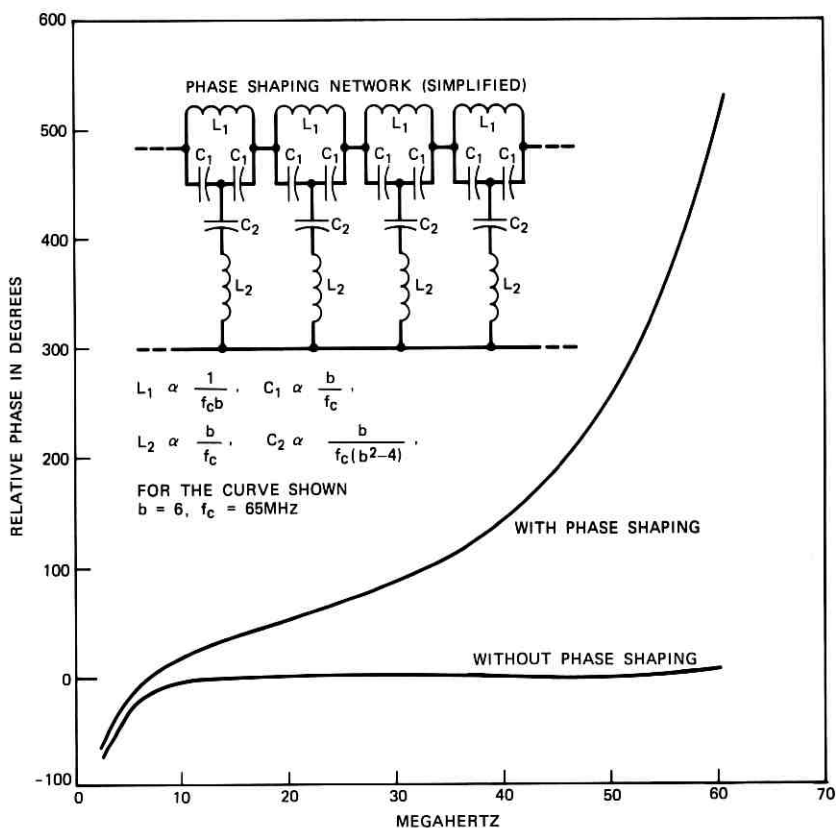
Fig. 9—Relative phase (linear component subtracted) of L5 regulating section—with and without phase-shaping networks in regulating repeater. The parameter b is a measure of the "stiffness" of the ϕ-f characteristic of the networks, and f_c is the frequency at which the phase shift of each of the four sections is 180 degrees.

IV. REPEATERED LINE DESIGN

As mentioned previously, the design of the repeatered line of a long-haul coaxial system centers largely about the equalization of the 4000-mile cable loss, and the cable used in the L5 system and its transmission characteristics are described next.

4.1 Coaxial medium

The typical L5 cable (new installation) will be a Coax-22 and will contain 22 disc-insulated coaxials and 42 polyethylene-insulated

conductor pairs. (A cutaway view of a Coax-20 is shown in Fig. 10.) The paired conductors in these cables are used for support systems such as order wire, fault location, and air-pressure telemetry.

The loss characteristic of the cable over the band of interest in L5 can be approximated by

$$\alpha(f, T) = 3.9002 \left(1 + \frac{0.0062}{\sqrt{f}} \right) \sqrt{f} + 0.0047f$$
$$+ 0.0043(T - T_o)\sqrt{f} \quad \text{dB/mile}, \quad (1)$$

where f is in MHz, T in °F, and $T_o = 55°F$.

The first term of the right-hand side is a result of the resistance loss of the copper conductors and, at L5 frequencies, owing to skin effect, this loss is proportional to the square root of frequency. The second term reflects the dissipation in the polyethylene disc spacers that maintain the relative positions of the inner and outer conductors, and it increases linearly with frequency. (This term is often called the "power factor" term.) The third term reflects the temperature dependence of the copper resistivity, and is also proportional to the square root of frequency.

Although the cable loss representation in eq. (1) is perfectly adequate over the transmission band of L5 and its predecessors, distortions in the cable loss will be encountered as system bandwidths are extended further, which in some cases may limit those bandwidths. For example, structural return loss (SRL) spikes are encountered, which result from the cable-stranding operation during manufacture. The strander periodically deforms the coaxials as they precess about the cable core, and, for a fixed stranding lay, the SRL spikes are coherent and result in relatively narrow-band transmission deviations from the behavior predicted by eq. (1). The current lay length for the outer coaxials in Coax-20 (cables having 20 coaxial tubes within the sheath) and Coax-22 is 36 inches $\pm \frac{1}{2}$ inch and results in SRL spikes at 157 $\pm$ 2.5 MHz. Other distortions are caused by seam interaction or "back twist," which is also a product of the stranding operation. These occur in the outer coaxials of the Coax-20 and Coax-22 around 170 MHz. The inner units of Coax-20 and -22 (8 in each case) and all the units of Coax-12, since they have shorter lay lengths, exhibit these effects at higher frequencies than those noted for the 36-inch lay length. For the purpose of designing the hierarchy of repeaters and equalizers that make up the repeatered line, the loss described by eq. (1) becomes the basic objective for repeater gain per mile of system, as discussed next.
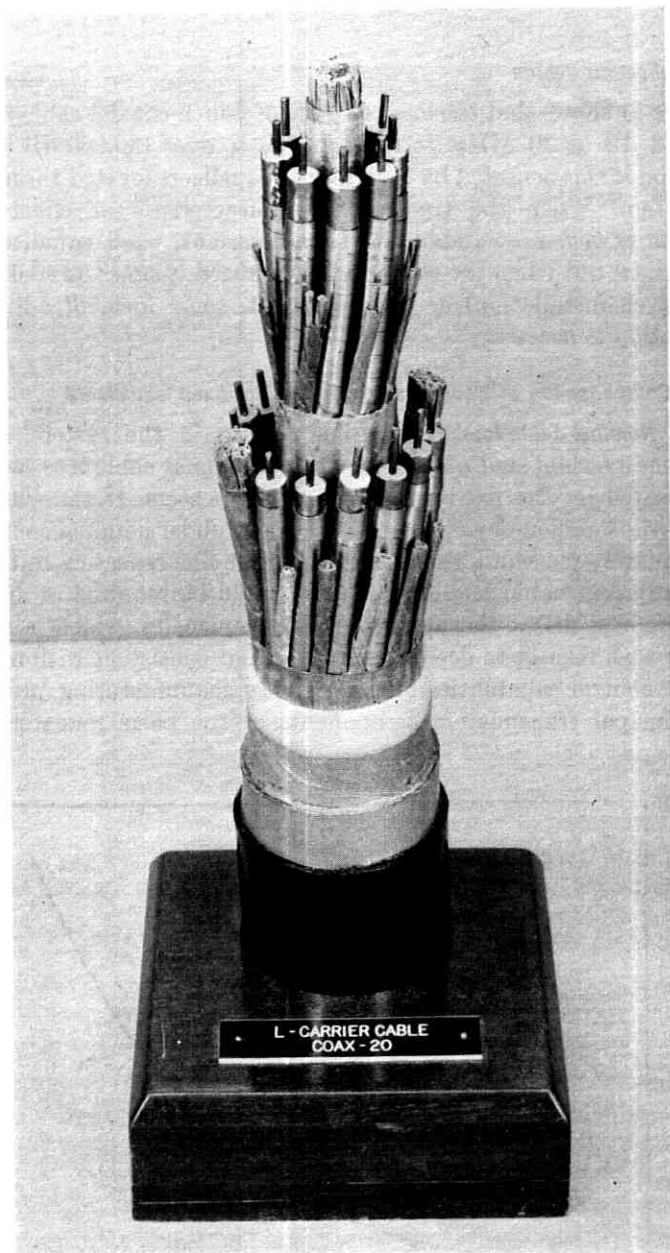
Fig. 10—Coax-20 of serrated seam construction. This cable is the backbone of nearly all existing L4 routes, and includes 52 paper-insulated wire pairs in addition to the 20 coaxials. New L5 construction will normally utilize Coax-22, specifically developed for L5.

4.2 Equalization design

Figure 11 shows that the loss of 1 mile of $\frac{3}{8}$-inch coaxial cable reaches about 33 dB at 70 MHz. It accumulates to over 130,000 dB in 4000 miles, and is compensated by a variety of equalizers located throughout the system. When the transmission characteristic or effect being equalized is *both* predictable *and* time-invariant, fixed equalizers are appropriate, but when the effect to be equalized is *either* unpredictable (in a given installation) *or* time-invariant, some form of adjustable equalization is necessary.

4.2.1 Compensation of transmission effects with fixed equalizers

4.2.1.1 Nominal cable loss. The principal loss in the system that is *both* time-invariant *and* predictable is the nominal cable loss, which is compensated by the fixed-gain basic line repeaters. If the 4000-mile system transmission objective were to be met under nominal conditions by adequately matching the gain of the basic line repeaters to the loss of the adjacent cable section, a match would be required of approximately 0.0008 dB at the uppermost L5 frequencies, which is neither realistic with respect to design complexity nor consistent with measuring and control capabilities of present-day manufacturing processes. The principal transmission components of the basic repeater are a
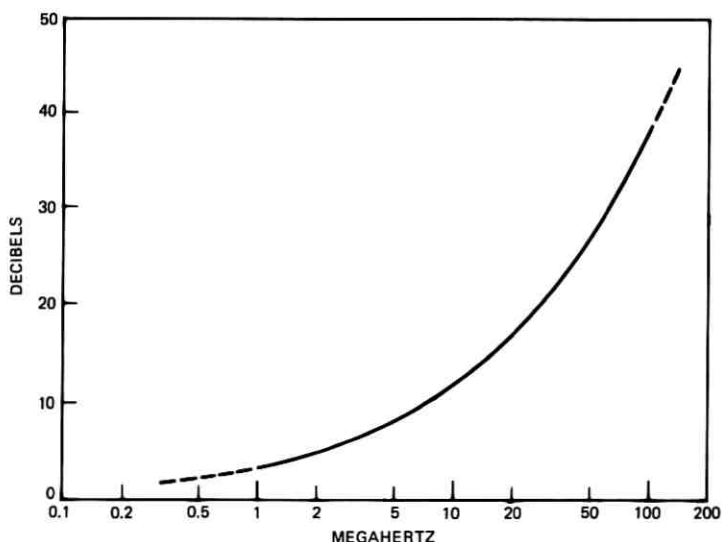

Fig. 11—Nominal loss of 1 mile of standard $\frac{3}{8}$-inch coaxial cable at 55°F.

low-noise preamplifier, a highly linear power amplifier, and an optional artificial line called a line-build-out network (LBO). These and the remaining components of the basic repeater are described in detail in Ref. 1.

4.2.1.2 Deviations from nominal repeater spacing. When the repeaters cannot be installed at the nominal 1-mile intervals, the resultant spacing deviations are compensated for by the abovementioned LBO networks which simulate the loss of a length of the coaxial cable, and are available in 0.1-mile increments from 0.1 to 0.5 mile. Because of obstacles, access problems, or other geographical limitations, it is usually impossible to achieve exactly 1-mile spacing between adjacent repeater sites. In such cases, the line repeaters are assigned one of the available LBO's to increase the loss of shorter cable sections to that of 1 mile of cable. For unusual obstacles such as river crossings, for example, a limited number of long repeater spacings may be permitted by requiring the repeater sections on both sides of the long section to be shorter than nominal. The administrative rules for all these cases are described in detail in engineering documentation available to the telephone company engineer at the time of route layout and route planning.

4.2.1.3 Average design error. In previous coaxial systems, such as L3 and L4, the equalization of the average difference between line repeater gain and cable loss (i.e., the average "design error") has been attempted with families of fixed equalizers called "design deviation equalizers" or simply "deviation equalizers." The functions of such equalizers are, first, to contain the transmitted signal levels at close to nominal values and, second, to enhance, in conjunction with all the other equalizers in the system, the ultimate transmission response of the system. While the first of these can often be guaranteed with relatively high confidence, the enhancement of the equalized transmission response is more difficult to achieve as a product of such equalization strategy.[1] This section discusses some reasons for this and some considerations involved in seeking an optimum overall set of equalizers (fixed plus adjustable).

As mentioned before, the initial design objective for a deviation equalizer is the complement of the average design error of the line repeaters. As used here, this error is the difference, on the average, between the gain of the line repeaters and the loss of the associated nominal length of coaxial cable. Ideally, the basis for computing the average difference would be the universe of all manufactured line

repeaters and cable. In reality, the basis must be the largest sample of each that is available before final manufacturing commitments must be made on the equalizers.

Even if the optimum information were available when desired, network realization limitations leave a residual error requiring further equalization, and when the deviation equalizers consist of a family of several different designs (as in L4, for example) there are several different realization errors. Naturally, the specific errors for different physical models of the "same" network also tend to be different (because of what is usually called the manufacturing deviation). Finally, even if the realizations were ideal, the average design error characteristic of the repeaters used as a basis for the deviation equalizer design will not always be characteristic of the set of repeaters in any particular line section.

The collective impact of these factors is a tendency to introduce "ripples" into the frequency response of the channel. Even though characterized by smaller amplitude deviations than the unrippled deviation, the rippled frequency response can often be more difficult to equalize. Of equal importance, the variations from line to line in the rippled characteristic resulting from the above factors will often be more difficult to track with a selected set of adjustable equalizers than would the line-to-line differences in the unrippled deviation (i.e., that existing before the application of the deviation equalizers).

To the extent that these factors contribute significantly to the ultimate residual error of an equalized channel, the transmission response of that channel would apparently benefit from minimizing the number of different types in the family of networks, and/or minimizing the number of locations at which the deviation equalizers are inserted. The straightforward application of these guidelines leads to a single deviation equalizer design inserted as infrequently (along the repeatered line) as possible, while the most extreme interpretation would lead to *no* deviation equalizers as such, a condition that probably could be achieved only with the addition of relatively costly adjustable equalizer stations, since the signal levels must still be contained within the repeater load capacity as the signal traverses the repeatered line.*

It is possible that the latter might be a desirable solution for a *single* equalizing section, which in L5 will be up to two power-feed spans.

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* The cost of a fully equipped L5 manhole equalizing station in a medium-cost zone is about twice that of a fully equipped regulating repeater station, which would normally be used otherwise.

However, when there is no attempt whatever (such as with deviation equalizers) to reduce the average misalignment over all equalizing sections to a value nearer zero, then all the adjustable equalizers in a long circuit would tend to be set to the same side of their nominal setting. As a result, there would be no tendency whatever to cancel (or randomize) the errors and transmission distortions originating in the equalizers themselves and in their inherent equalizing limitations, and these would accumulate systematically in a long circuit, quite possibly becoming the major source of residual overall transmission deviation. Short of this extreme, a solution based on a single network design and application in a minimum number of locations is attractive from several viewpoints, and such a solution has been incorporated into the L5 design.

The L5 deviation equalizer design provides one-third the total "design error" equalization required by a 66-mile repeatered line (approximate average power span length) in single networks located in the transmitting and receiving main stations and at the midspan equalizing station. The network has been designated the "22X" deviation equalizer, since the design objective is approximately the design error of 22 miles of L5 repeatered line. The particular method used to realize the deviation equalizer design is described in detail in Ref. 1. As described there, the L5 deviation equalizer, while in principle correcting the average design error, is realized in the final design stages using computer optimization routines that interact with the available adjustable equalizers. Figure 12 shows the signal misalignment as a function of distance, per 3 dB of 66-mile correction, employing the 22X approach and assuming uniform accumulation of the misalignment. As far as signal-to-noise penalty is concerned, it can be seen that this arrangement is equivalent to pre- and postequalization, in equal
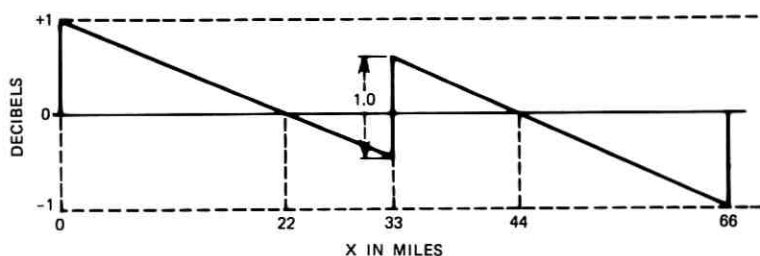

Fig. 12—Equalization of the average design error. The relative signal level is shown for a case involving 3 dB of misalignment in 66 miles; three identical deviation equalizers are placed at the transmitter and the receiver, and midway between.

Table III — Deviation equalizer application rule

X in miles	Power Span of X Miles		
	TRMTG RPTR	MIDSPAN RPTR	RCVG RPTR
55 to 75	Yes	Yes	Yes
33 to 55	Yes	No*	Yes
11 to 33	No	*	Yes
Less than 11	No	*	No

* No midspan equalizing repeater required for $X < 37.5$ miles.

parts, of 44-mile sections having ±1-dB signal excursion in tandem with 22-mile sections having ±0.5-dB signal excursion, each section incurring signal-to-noise penalties very near zero (see Fig. 3).

Naturally, administrative rules are required to assist the telephone company engineer in the determination of his 22X equalizer needs. With "quantum steps" of 22 miles of fixed equalization, it is possible to specify the fixed equalization to within $\pm22/2$ miles of the ideal value, and the application rules specified for L5 are shown in Table III. Note that execution of the tabulated rules should approach compensation, on the average, of the design error in a long (multispan) system that operates at a uniform mean temperature. In so doing, the rules should result in slight over- or undercompensation of the individual equalizing sections which will be of varying lengths. This effect, along with the significant variations in mean cable temperature that normally will be characteristic of long systems, should result in a tendency to randomize the adjustable equalizer settings, thereby mitigating the systematic accumulation of the errors and distortions originating in the adjustable equalizers themselves, and in their inherent equalizing limitations.

4.2.2 Compensation of transmission effects requiring adjustable equalizers

As discussed above, those transmission effects that are *either* time-variant *or* unpredictable (or both) require some form of adjustable equalization. Adjustable equalizers can be automatically or manually adjusted and the process can be carried out either in- or out-of-service, as particular circumstances and considerations warrant. In the case of L5, each of these possible types of equalizer and equalization processes is utilized among the various members of the equalization hierarchy.

4.2.2.1 Cable-temperature effect. The largest time-variant transmission deviation in a coaxial cable system, normally orders of magnitude greater than any other, is the cable-temperature effect, shown in Fig. 13 for 1 mile of ⅜-inch cable and for a 20°F change in cable temperature. Extrapolated to 4000 miles, the temperature dependence depicted here corresponds to about ±3000 dB of loss variation at the uppermost L5 frequencies, uniformly distributed over that distance. In L5, this is corrected at intervals not greater than 7 miles in "regulating repeaters" which, in addition to compensating for the nominal cable loss, provide automatic equalization of the cable-temperature effect. The regulating repeater compensates for the changing cable loss within a "regulating section" which, in typical installations, has an average length of approximately 6 miles.

In addition to the components found in a basic repeater, the major transmission components of the regulating repeater are a pair of adjustable equalizers having the adjustable loss frequency capability shown in Fig. 13 for the cable-temperature effect. The first equalizer operates in a closed-loop feedback mode and postequalizes for a portion of the preceding regulating section. The second equalizer operates in an open-loop mode by sensing the resistance of a thermistor buried in the ground near the cable, thereby pre-equalizing for a portion of the cable-temperature effect of the ensuing regulating section. These two equalizers restrain the maximum level deviations resulting from the
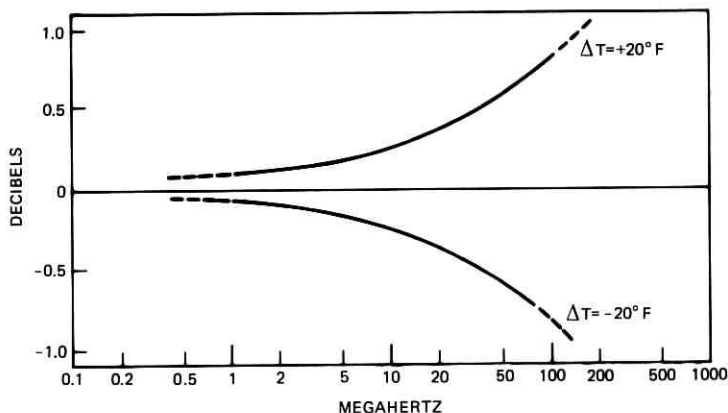

Fig. 13—Effect of changing earth temperature on 1-mile cable loss. The indicated temperature range is only rarely exceeded within the continental U. S. at the cable burial depth of 4 feet.

cable-temperature effect to about ±2.5 dB within a regulating section through the use of the pre- and postequalization method. When the system is properly aligned, the pre-equalizer in one regulating repeater and the postequalizer in the subsequent regulating repeater each compensate for one-half the cable loss variation of the intervening cable section. The remaining components and a more detailed description of the regulating repeater operation are included in Ref. 1.

4.2.2.2 Repeater-temperature effect. The other major time-variant transmission deviation is the repeater-temperature effect. This results from both the regulating repeater tracking error associated with equalizing the cable-temperature effect (a maximum of 0.10 dB per regulating section) and the direct effect of changing temperature on the transmission responses of the various line repeaters (less than 0.02 dB per repeater section). Suitable allowances had to be made in the repeater signal loading requirement to permit this effect to be equalized at the receiving end only of each power-feed span (up to 75 miles long). This is accomplished in the "dynamic equalizer," a cause-associated, adaptive equalizer with memory,* the important characteristics and operational features of which are described in Ref. 1.

4.2.2.3 Statistical manufacturing deviation. The largest time-invariant transmission deviation requiring adjustable equalization is the statistical manufacturing deviation that results from the (usually small) differences in electronic component values within their specified tolerance ranges. Thus, the exact frequency shape of this deviation is not predictable for any given collection of repeaters in a particular installation, although it can be predicted to fall within certain overall limits. In L5, this deviation is compensated by equalizers E1 and E2 which are manually adjustable because the deviation is time-invariant once identified. These equalizers consist of 28 adjustable Bode-type equalizers ("bump" equalizers)—18 in the E2 and 10 in the E1. Both E2 and E1 equalizers are located in the transmitting and receiving main stations, while only the E1 equalizer, mechanically adapted for manhole use, is utilized at the midspan equalizing repeater. The deployment chosen for the equalizing stations is the result of seeking a reasonable balance between the amount of misalignment that would have to be tolerated by the line repeaters, with resultant signal-to-noise penalties and effects on repeater power requirements,

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* The memory in the dynamic equalizer permits the transmission response of the equalizer to remain unchanged if any equalizing pilots are lost.

and the proliferation of the relatively complex (and relatively costly) equalizing repeaters that would otherwise be required to keep the misalignment and associated signal-to-noise penalties small.

To estimate the static misalignments to be encountered, it is necessary to know some transmission properties of the repeaters which, at the outset of a system design, either are in the form of repeater requirements or, at times, are based on previous experiences with similar transmission components. In the case of the L5 design, a requirement was imposed on the line repeaters that (i) the average gain of the repeaters match the nominal cable loss within ±0.1 dB and (ii) the spread among the gains of all manufactured repeaters would be less than ±0.1 dB with respect to that average. This, in combination with the deviation equalizer strategy described in the preceding section, made it possible to specify and design a single midspan (between power-feed stations) adjustable equalizer having both adequate range and adequate margin (against the possibility, for example, that the repeater transmission requirements turned out to be unrealistic, or too costly to achieve, and might thus have been relaxed somewhat). As mentioned earlier, this also resulted in the spacing of the relatively expensive power-feed stations being limited by line power considerations, rather than by equalization and repeater load capacity considerations.

The selection of the type of equalizer terms, or "shapes," to be used and the number of terms in the equalizers, as well as the deployment of the shapes in the frequency domain, was based in part on computer simulations of the L5 channel as it was estimated to be, and in part on engineering judgments that evolved from experiences with earlier systems such as L3 and L4. For example, it was judged possible to meet the line repeater-cable loss matching requirement with about four or five peaks in the matching error as a function of frequency. (Five was the actual number.) It was further reasoned that (i) the action of the deviation equalizers, intended to complement the line repeater error function, would tend to double that number of peaks in the resultant "fixed equalized" characteristic; (ii) the midspan equalizer, while as simple as possible, should have a number of terms equivalent to the number of deviation peaks expected in the characteristic for which it was to compensate (8 to 10); (iii) the action of the midspan equalizer would tend to increase the number of deviation peaks in the resultant "coarse-equalized" characteristic by about the number of terms in the equalizer (presumably, 8 to 10). This leads to a main station equalizer being required to deal with a transmission characteristic

having potentially 16 to 20 peaks, whereupon the equalizer would correspondingly be required to have a similar number of terms. This type of reasoning provided starting conditions for various computer-aided optimization routines that ultimately resulted in the 10-term E1 equalizer and the 18-term E2 equalizer. Similar computer simulations were utilized to evaluate the effectiveness of such different types of equalizer shapes as the Bode-equalizer-derived "bump" shapes (as used in L4) and cosine equalizers (as used in L3). In each simulation undertaken, a given number of terms could be deployed more effectively with "bumps" than with alternatives, consequently leading to the selection of the "bump" equalizers for L5. While the initial analyses and selection of equalizer characteristics had to be based on *expected* L5 channel performance, the *final* deployment of "bump" shapes, as manufactured for the initial commercial application, was based on actual L5 repeatered line realizations in the L5 field trial. (The field trial took place between Cedarbrook and Netcong, N. J., from mid-1970 to mid-1972, with each of four coaxial lines equipped and evaluated over a distance of approximately 100 miles. The trial resulted in a variety of important system improvements.)

The manually adjustable equalizers are intended to be set upon initial installation and only rarely thereafter—for example, when a substantial increase in the amount of equipment operating in the manholes results in significant changes in the annual mean operating temperature of the repeaters. Readjustments of the E1 and E2 equalizers under such conditions will restore the mean-annual setting of the dynamic equalizers (E3) to midrange, will improve the overall transmission response (since the 28-term E1/E2 combination is more effective at equalizing the temperature offset effects than the four-term dynamic equalizer E3), and will improve signal-to-noise performance by reducing the maximum misalignment. Such readjustment would be accomplished out-of-service and would involve determination of the channel misalignment at a set of frequencies (three per adjustable term) selected to permit optimum channel equalization in a mean-squared-error sense. The misalignment of the channel is measured using programmable transmission measuring equipment under the control of what amounts to a special-purpose computer designated the Equalizer Adjustment Unit. The detailed implementation of equalizer adjustment is described in Ref. 1.

4.2.2.4 Other factors affecting system transmission. For completeness, two other factors that result in time-variant transmission effects are

mentioned here. One of these is the aging of the elements making up the system. Aging effects usually are much smaller and much slower than any effects already discussed. In a solid-state system designed with careful attention to component selection and component stresses, there is no aging mechanism comparable to that which was characteristic of the electron tube systems L1 and L3. In L5, aging effects are expected to be small enough to be insignificant with respect to transmission quality or with respect to such operational aspects as equalizer adjustment intervals.

The second of these effects is that associated with maintenance which, along with the aging effect, is normally one of the smallest time-dependent sources of change in the transmission response of a system. Over long periods of time there will inevitably be occasional repeater failures with subsequent replacement of these repeaters by spares not having the same transmission response as the failed repeater. (Present estimates of repeater reliability suggest a mean-time-to-failure in a maximum-length power span of 75 miles because of repeater failure at between one and two years. Individual failures are, of course, protected by the LPSS.) When they become large enough to be detected, both aging and maintenance effects would normally be compensated by one of the occasional readjustments of state equalizers E1 and E2.

4.3 Comparison of L5 features with some predecessors

When the characteristics of the L5 repeatered line are examined, it is evident that it shares many features with previous systems, and thus in some ways is a hybrid offspring of the L3 and L4 predecessors. For example, while the adjustable equalizers E1 and E2 consist of bump shapes like L4, the adjustments are made out-of-service and in such a way as to minimize the residual mean-squared error of the equalized channel, like L3. The cable-temperature effect is compensated by using both open- and closed-loop regulators which pre- and postequalize the changes in cable loss in equal parts, like L4. But the 22X deviation equalizer strategy of L5 is much more akin to the L3 approach. (In L4 there are five different designs, one of which is installed in every regulating repeater.) While all three systems utilize line-build-out netowrks to provide some spacing flexibility, they have fault-locating systems that are quite different. And only the L5 system supplements the basic fault-location system with a centralized, computer-operated transmission surveillance system

capable of measuring and monitoring a wide variety of system operating parameters.

V. CURRENT STATUS OF COAXIAL SYSTEM DEVELOPMENT

At the time of this writing, the development of the L5 system in its initial configuration had been completed with all important noise and transmission objectives having been met. Continuing efforts in the software offerings of the transmission surveillance system are expected to increase the already considerable capabilities for automatic and semiautomatic system maintainance, fault location, and trouble diagnosis. A modest level of continuing effort involving the optimization of deviation equalizer designs and modification of equalizer deployment should further improve the system transmission response and reduce system noise.

Effort is also under way to replace the existing mastergroup multiplex equipment (MMX-2c) with individual mastergroup translators, resulting in reduced noise contribution from that level of multiplex at lower cost. The possibility of alternative protection strategies for the jumbogroup level of multiplex are also being examined. In addition, the basic frequency allocation described in this paper is being reviewed with an eye toward possible spectrum reorganizations that would permit an increase in L5 capacity to 22 mastergroups using a new family of mastergroups and multimastergroup multiplex equipment.

Finally, questions continue to be raised with respect to future coaxial systems—whether or what or when they may be. Although advanced coaxial systems having substantially greater capacity than L5 seem clearly to be within the technological capabilities of the relatively near future, the need and economic desirability of such systems has not yet been established. Whether or not L5 becomes the last of the major long-haul coaxial developments will be determined in the years ahead primarily by considerations of circuit growth patterns and economic conditions and by the relative viability and availability of alternative instrumentalities such as possible new microwave radio, millimeter waveguide, and optical fiber systems.

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L5 SYSTEM:

Repeatered Line

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

I. INTRODUCTION

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

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

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

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* A 2.1-percent factor has been included in this loss to account for miscellaneous factors such as stranding.
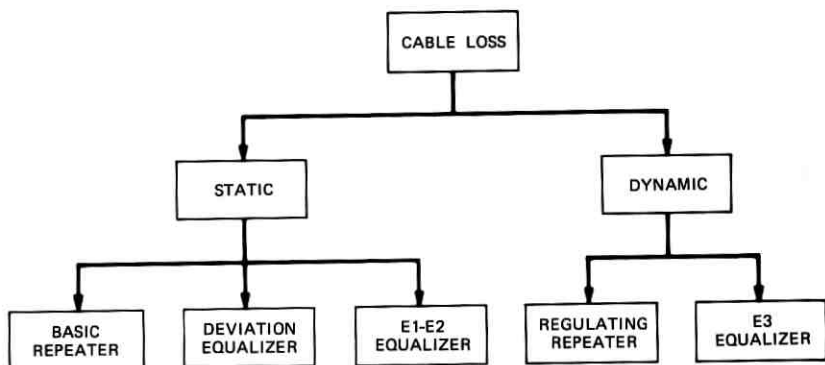
Fig. 1—Equalization strategy.

where the average values of the parameters are

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

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

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

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

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

1.1 Static equalization

Static equalization of the cable is provided in three levels. The first is provided by the basic repeater. With the repeaters spaced at 1-mile

intervals, the gain required to match the static component of the cable attenuation given in (1) varies from 4.97 dB at 1.6 MHz to 32.04 dB at 66 MHz.

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

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

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

1.2 Dynamic equalization

Equation (1) shows that the cable attenuation varies as the temperature of the cable changes. In the L5 system, the cable is buried four feet beneath the ground surface, where the maximum temperature deviation expected in the United States is $\pm20°F$. This results in a 52-dB change in cable attenuation at 66 MHz for 75 miles. Such change in attenuation is compensated for by repeaters that are called regulating repeaters. These repeaters automatically compensate for changes in the cable attenuation by sensing the temperature of the earth at cable depth and by detecting the level of a pilot tone which is inserted into the message signal at L5 main stations.

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

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

II. BASIC REPEATER

2.1 Basic repeater function

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

return loss, and temperature coefficient. The simultaneous achievement of all these specifications is a major accomplishment in the design of the L5 system.

2.2 Basic repeater design requirements

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

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

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

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

To summarize, the basic repeater objectives dictate a shaped-gain, low-noise, high-power ultralinear repeater. One method of achieving the shaped-gain requirement would be to design a flat-gain amplifier and a separate network for the shaped loss. The shaped loss would be equal to the maximum gain required minus the minimum gain required

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

2.3 Basic repeater realization

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

2.3.1 Twin jack and bridging pad

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

The bridging pad has an impedance of about 2000 ohms facing the twin jack. Since it connects with a coaxial plug, there is also a parasitic capacitance to ground. To maintain the 75-ohm impedance required at the repeater input and output, an inductor was designed into the twin jack. This is shown schematically in Fig. 3. By choosing L such that $\sqrt{L/C} = 75$ ohms, a lumped section of 75-ohm coaxial cable was approximated, and the impedances were maintained. The shunting effect of the 2000-ohm resistor is compensated for by the impedance of the low-frequency networks that are discussed in Section 2.33.
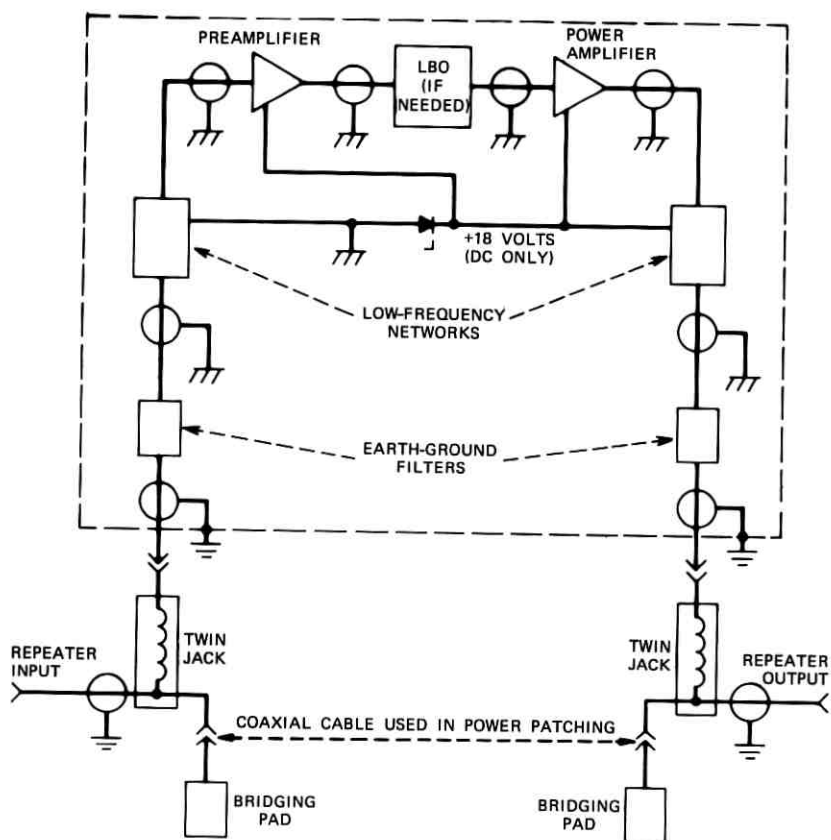
Fig. 2—Basic repeater.

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

2.3.2 Earth-ground filters

The earth-ground filters, shown at the repeater input and output, consist of sections of high-voltage capacitors and ferrite beads which form a filter. They provide transmission of signal and power through the outer wall of the repeater (which must be kept at earth ground for safety reasons) to an inner "floating ground." Signals developed across the high-voltage output capacitors are then sufficiently attenuated by the filter so as not to couple into the input high-voltage capacitor, and vice versa. Failure to provide this filtering action would result
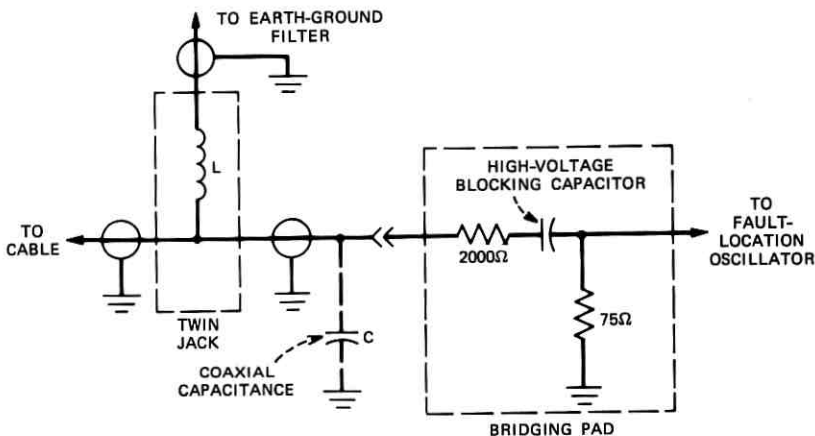
Fig. 3—Twin jack.

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

2.3.3 Low-frequency networks

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

G1 is a gas tube surge protector having a dc voltage breakdown of about 90 volts and, under normal circuit conditions, introduces only 1.5 pF of shunt capacitance. Inductor L1 provides for the separation of the dc line current from the L5 message signal. It maintains a high reactance over the frequency band to keep signal losses to a minimum.
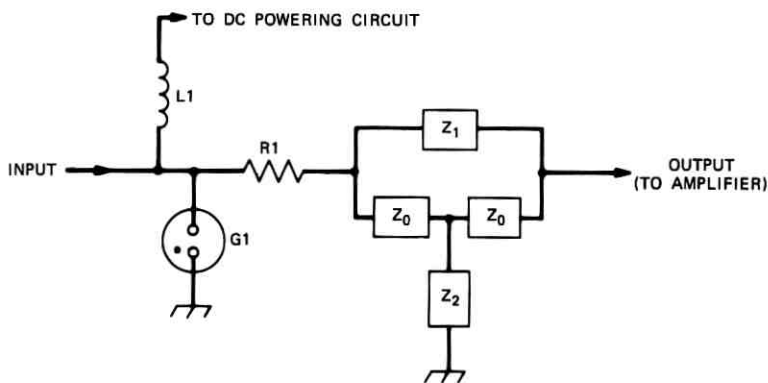
Fig. 4—Low-frequency network.

2.3.4 Line-build-out network

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

2.3.5 Amplifier design considerations

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

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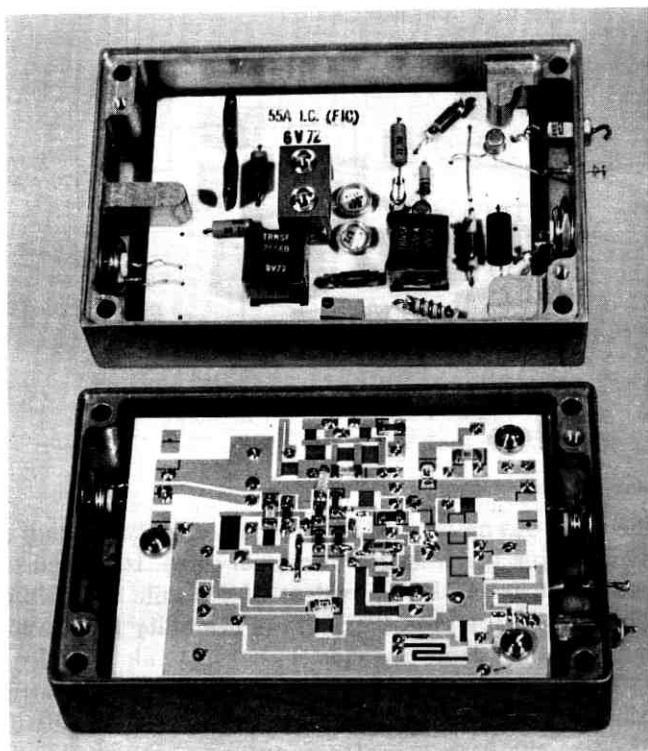
Fig. 5—Tri-level preamplifier.

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

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

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

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

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

Two adjustable elements are in the preamplifier—a capacitor and a potentiometer. Each has a very limited range and is used for factory adjustment of the gain response, particularly at high frequencies. Computer sensitivity runs were used to choose the variable elements for matching the normal amplifier gain shape variation. The capacitor affects the balance between loop and local feedback and hence influences the closed loop gain through the $\mu\beta$ effect,[4] which varies the repeater gain only at the very high frequencies. The potentiometer is in the bridged-T β-network. Changes in its resistance directly vary the β loss and, hence, the amplifier gain. This adjustment changes a broad, high-frequency shape and is made after the total repeater has been assembled, to set the gain at 42.880 MHz (the temperature pilot) to within ± 0.01 dB of the nominal value. This precise gain adjustment
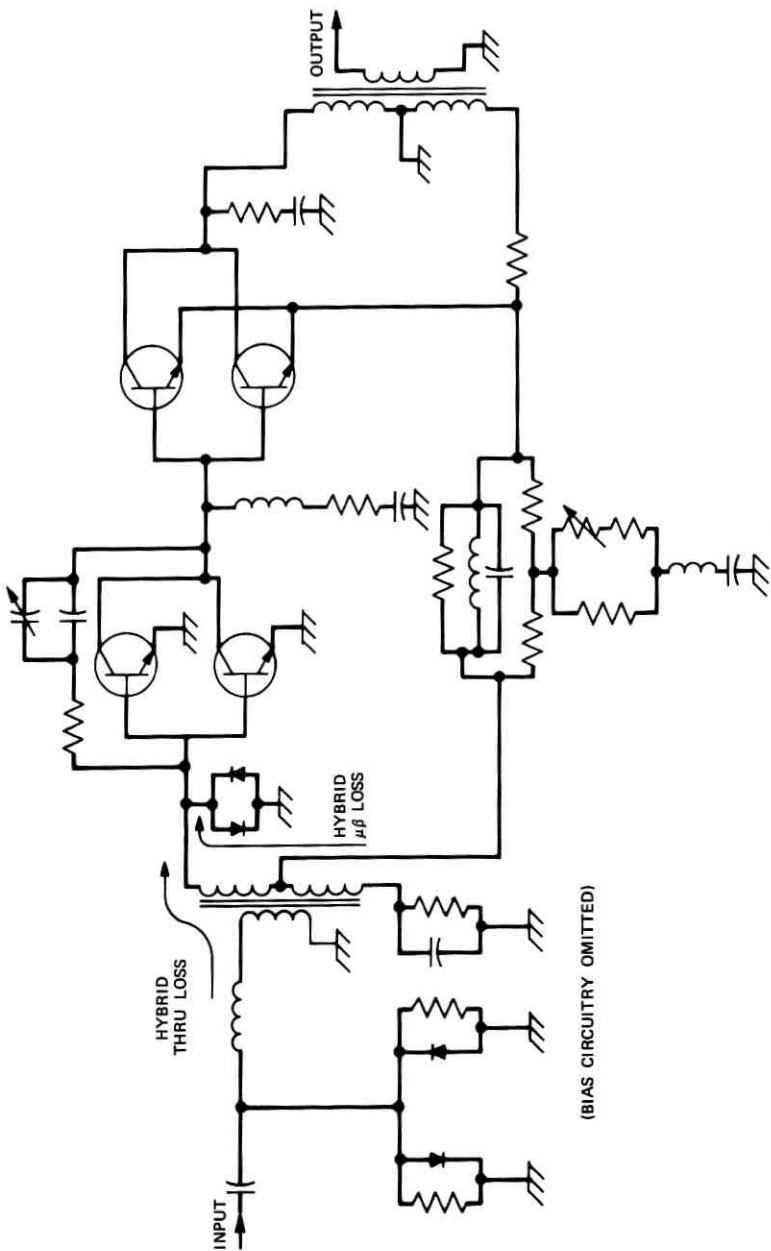
Fig. 6—Preamplifier schematic.

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

2.3.5.2 Power amplifier. The most difficult objective to meet in the basic repeater was the intermodulation distortion. After an extensive program of computer modeling of transistor and circuit nonlinearities[6] and laboratory evaluation, the circuit topology shown in Fig. 7 was chosen for the power amplifier. It has shunt feedback at the input and hybrid feedback at the output. The μ path is a common emitter-transformer-common-base arrangement. The interstage transformer provides current gain to minimize the effect of distortion in the common emitter stage and also provides a more nearly optimum set of interface impedances between the stages. An autotransformer was chosen as the output hybrid to obtain maximum bandwidth and minimum phase shift.

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

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

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

2.3.6 Surge protection

Buried cable systems, although shielded, are still subjected to a number of transients induced from within and without the system. These include (i) high-voltage line turnup,[6] (ii) power patching and subsequent insertion of an uncharged repeater into the high-voltage line (iii) lightning, and (iv) 60-Hz induction. To protect the repeater
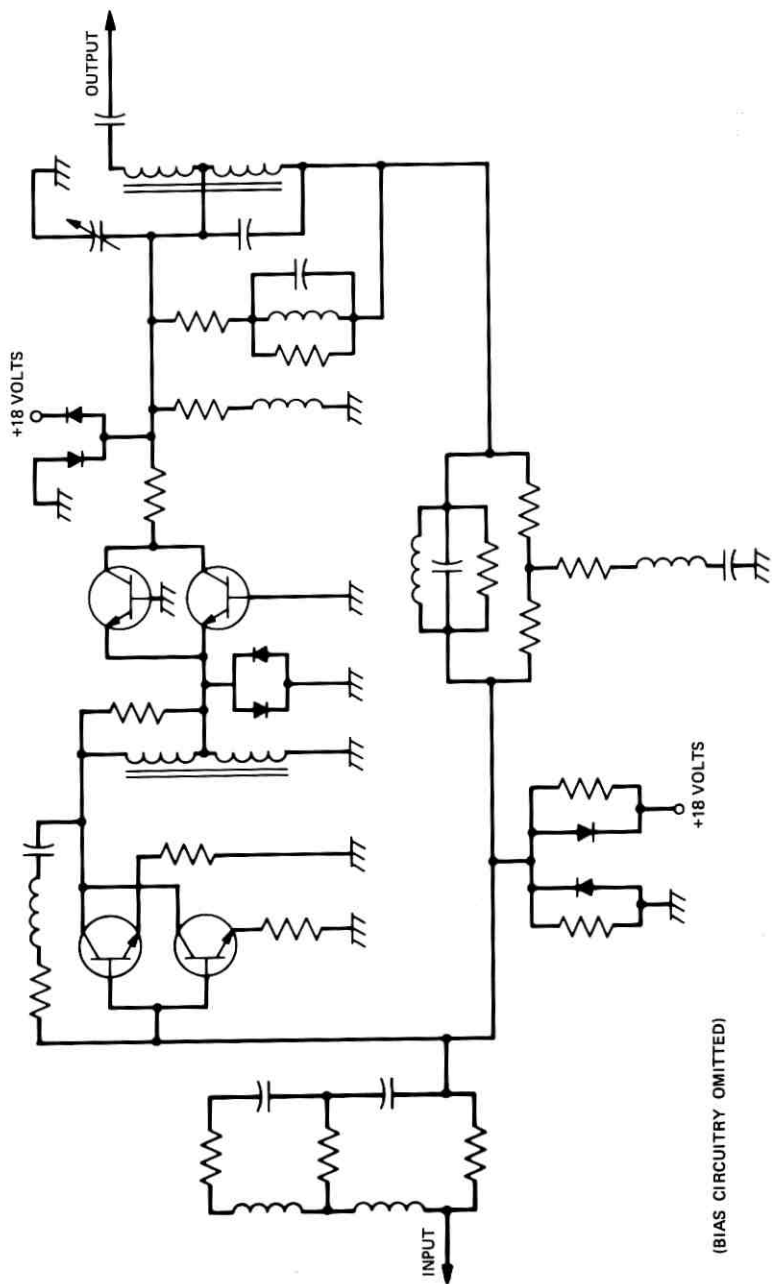
Fig. 7—Power amplifier schematic.

OUTPUT

+18 VOLTS

+18 VOLTS

INPUT

(BIAS CIRCUITRY OMITTED)

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

2.3.7 Basic repeater performance

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

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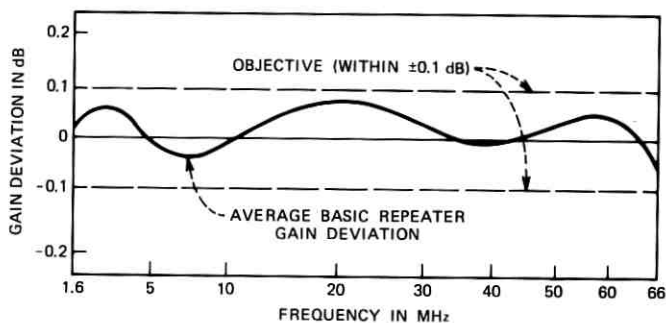

Fig. 8—Average deviation of basic repeater gain from the average loss of 1 mile of coax cable.
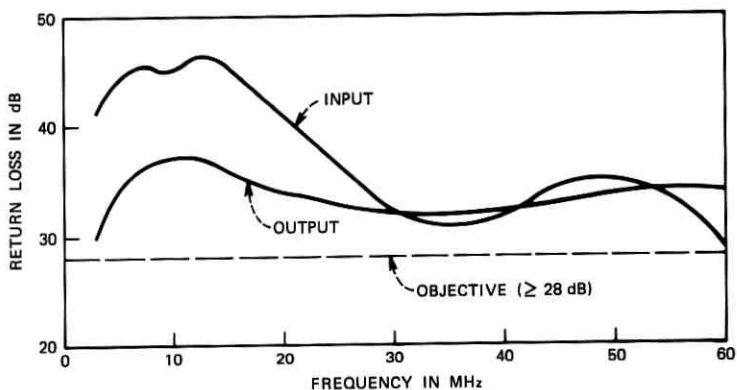
Fig. 9—Basic repeater return loss (average of 2071 production repeaters).

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

III. DEVIATION EQUALIZER

It has been pointed out that exact equalization of the static attenuation component of the cable loss is not feasible. The design transmission objective of the basic repeater is to match the nominal cable attenuation characteristic to within ±0.1 dB. The difference between
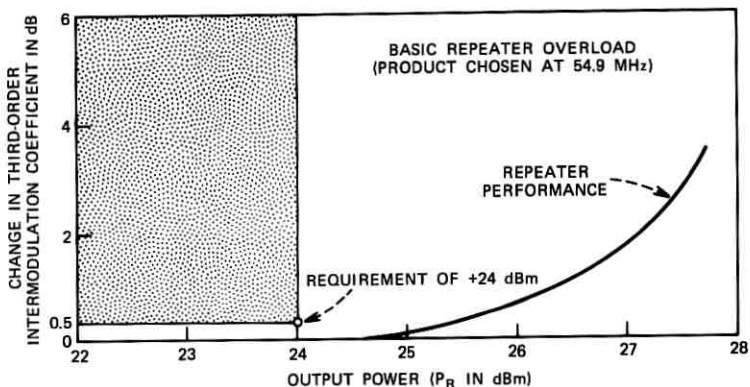
Fig. 10—Basic repeater overload (product chosen at 54.9 MHz).
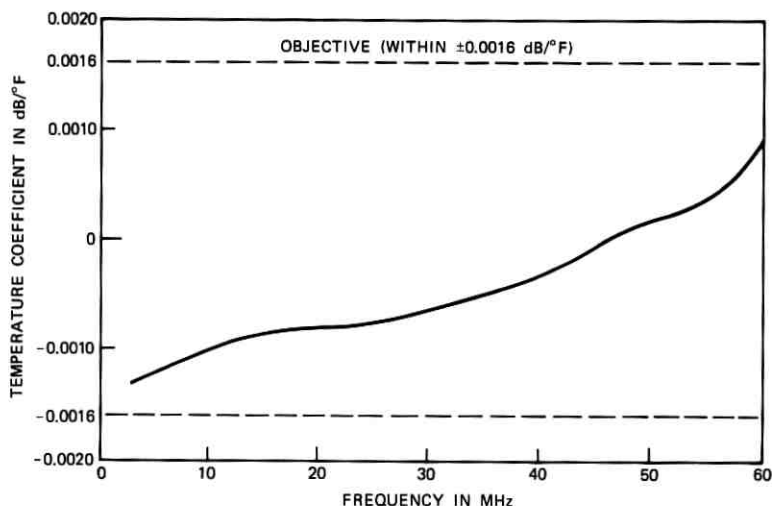
Fig. 11—Basic repeater temperature coefficient.

the gain of the average repeater and the nominal loss of the cable is called the average design error and is obviously predictable. Equalization of the average design error is accomplished by providing fixed equalizers in the transmitting and receiving repeater in the main stations and in the equalizing repeater located at approximately the midspan location between main stations. In previous L-carrier systems, the loss characteristic of the deviation equalizer was determined as
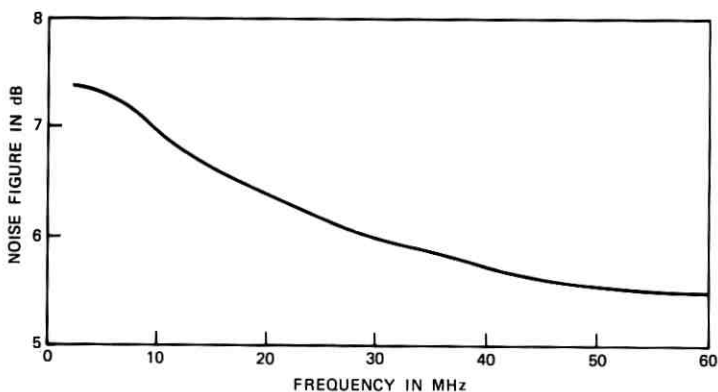
Fig. 12—Basic repeater noise figure.
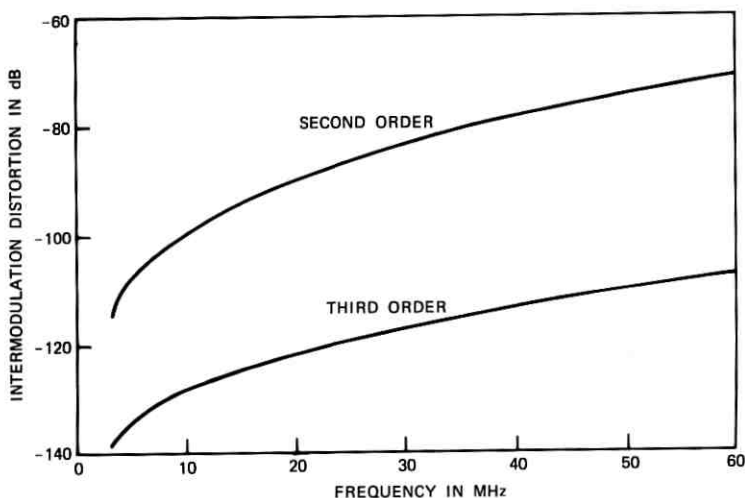
Fig. 13—Basic repeater intermodulation distortion.

follows:

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

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

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

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

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

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

within the capabilities of the variable equalizers provided in the system.

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

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

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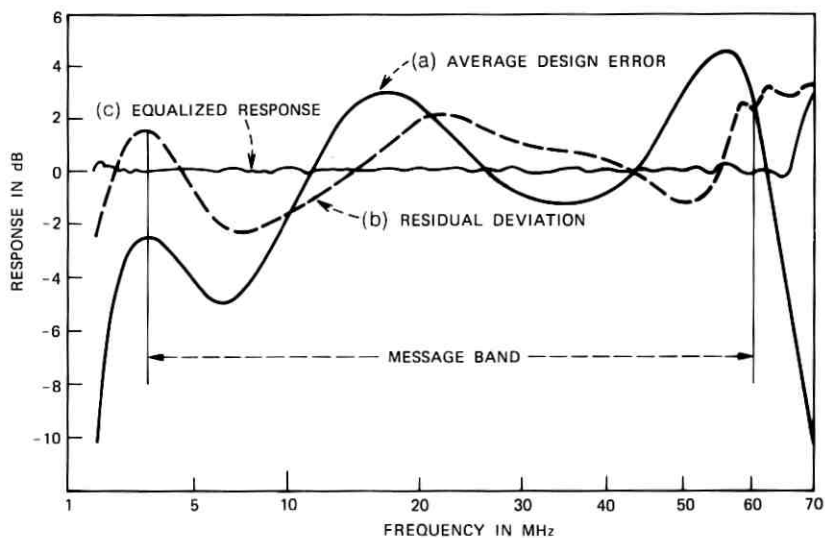

Fig. 14—Static equalization.

IV. ADJUSTABLE STATIC EQUALIZERS

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

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

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

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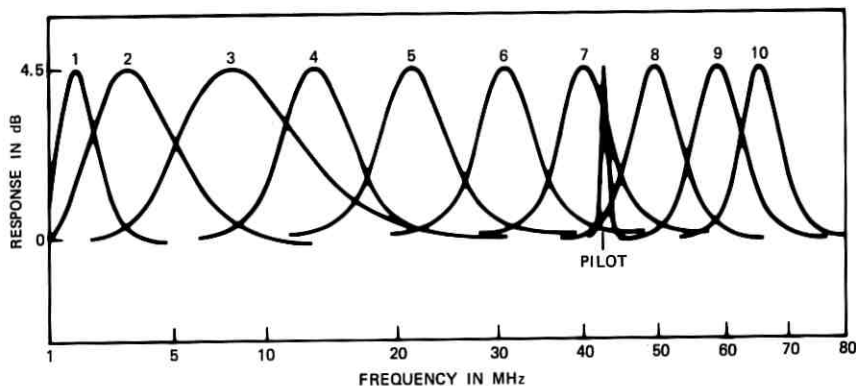

Fig. 15—E1 equalizer shapes.

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

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

4.1 E1 equalizer

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

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

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

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

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

The most difficult design problem in the E1 (and E2) equalizers was the achievement of the overall frequency response. The gain or loss of each amplifier or network comprising the equalizers is adjusted during its manufacture to be within ±0.05 dB of its respective nominal value. The requirement on the overall equalizer is that, in its reference state, the gain of the equalizer should be within ±0.1 dB of zero when all the Bode networks are in their flat condition. This has been achieved by providing a trimming network adjusted during the assembly of the E1 equalizer to trim the characteristic of the E1 to within the desired requirements. Because the network and amplifier gains are tightly controlled, the trim equalizer is a simple network containing only a flat and a slope term.
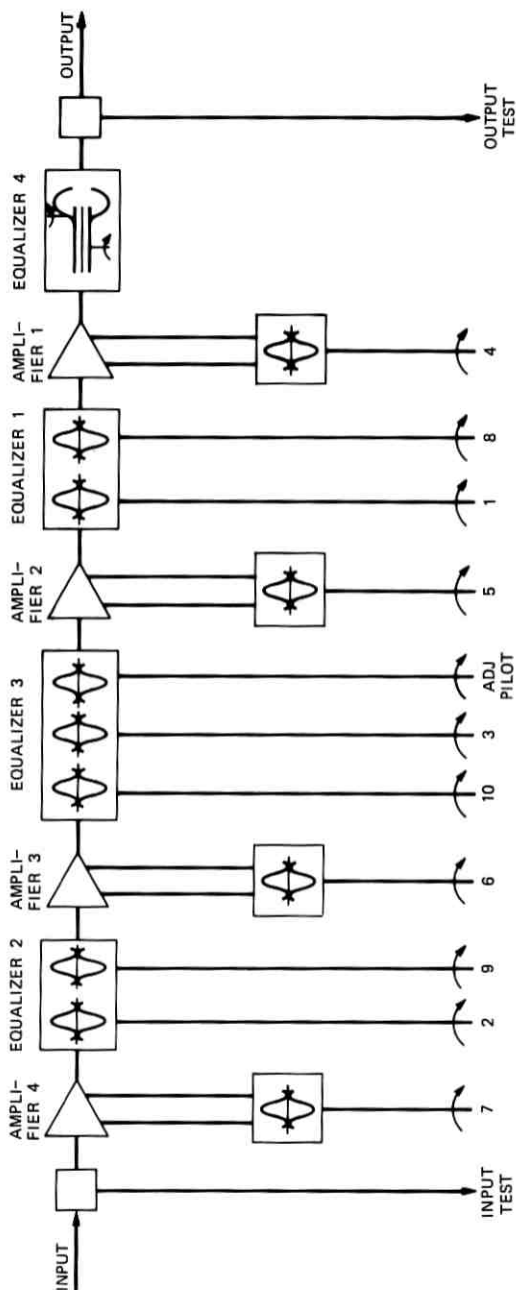
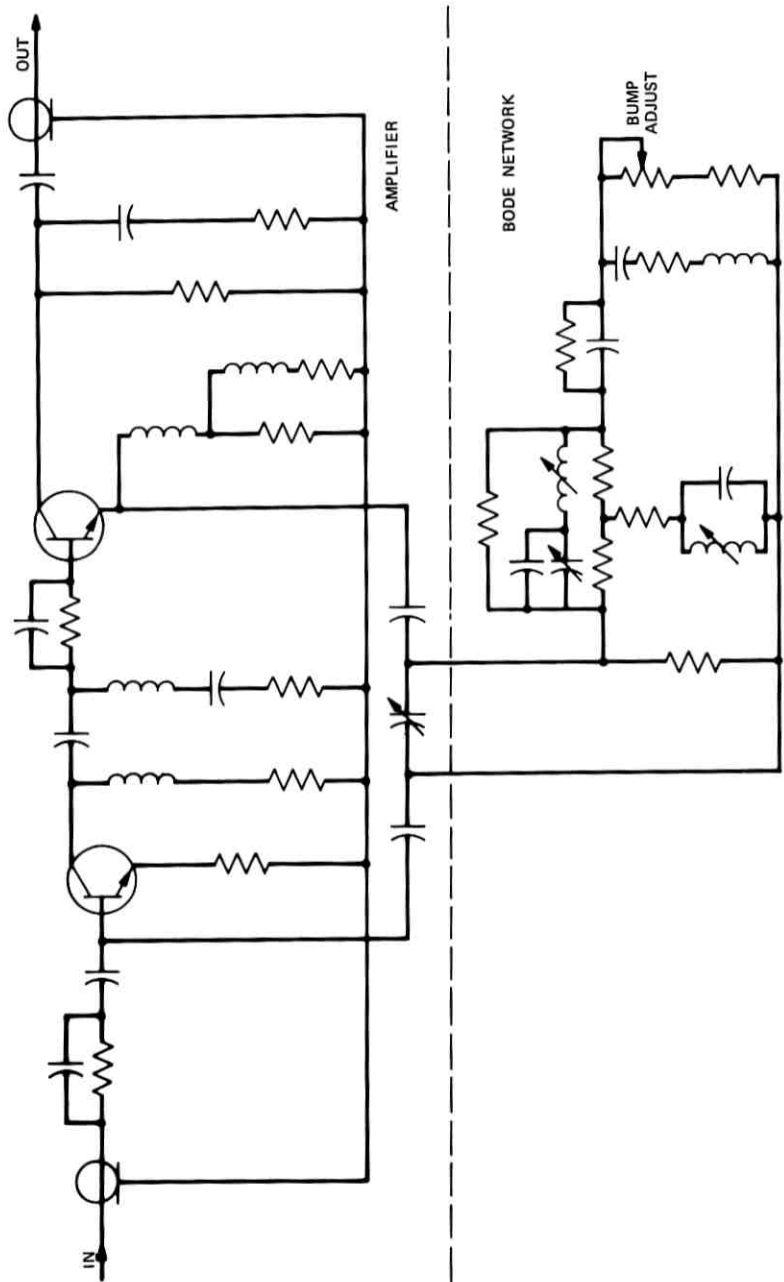Fig. 16—E1 equalizer transmission circuitry.
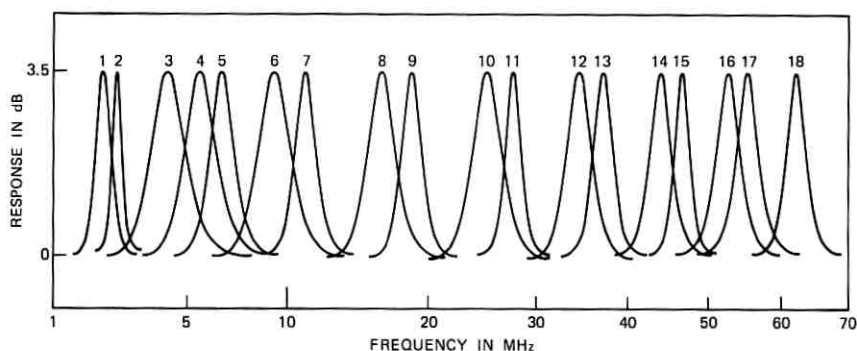
Fig. 17—Equalizing amplifier schematic.

Fig. 18—E2 equalizer shapes.

4.2 E2 equalizer

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

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

Table I — Static equalizer characteristics

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

4.3 Equalizer adjustment algorithm

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

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

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

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

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

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

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

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

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

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

$$G_k = \frac{\partial \text{MSE}}{\partial g_k} = -2\langle B_k(f), E(f)\rangle$$

$$= 2\langle B_k(f), \sum_{j=1}^{28} g_j B_j(f)\rangle - 2\langle B_k(f), M(f)\rangle \tag{5}$$

$$= 0 \text{ for all } k = 1, 2, 3, \cdots, 28,$$

where we define

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

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

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

where

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

and

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

The gain vector is

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

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

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

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

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

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

(iii) Repeat the above steps.

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

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

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

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

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

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

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

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

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

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

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

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

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

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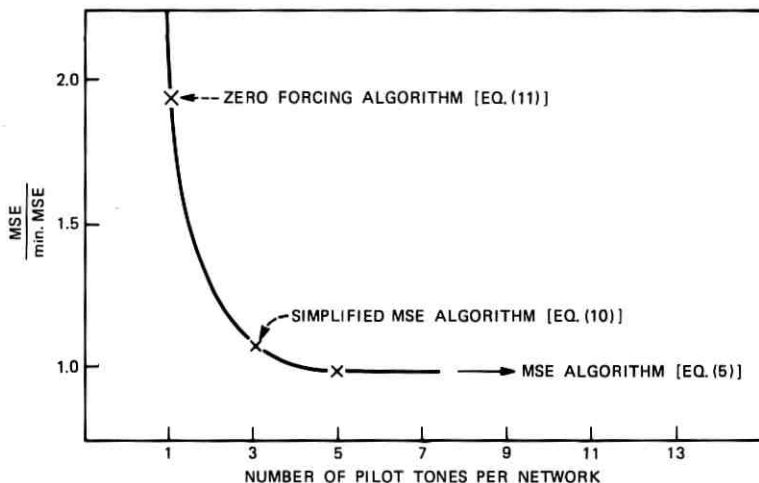
Fig. 19—MSE values by different algorithms.

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

4.4 Equalizer adjustment unit and equalization process

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

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

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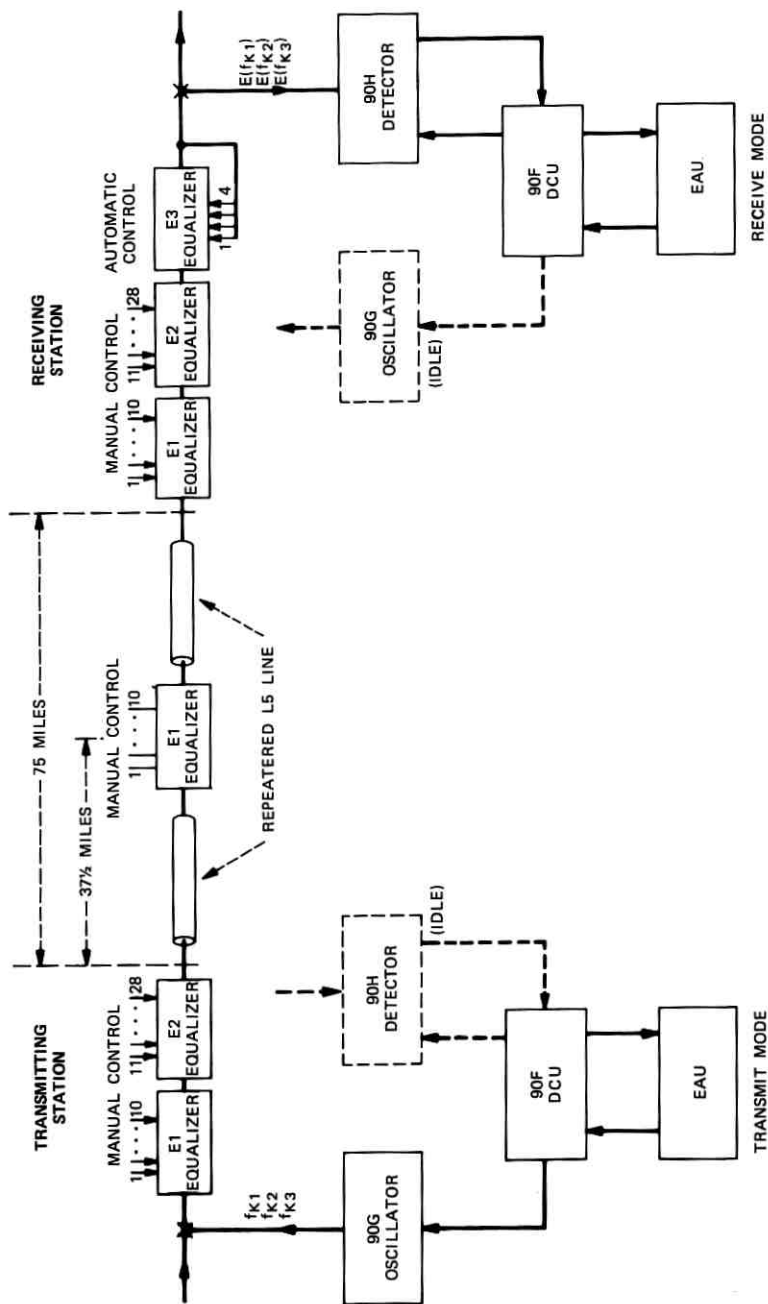
Fig. 20—Equalization plan.

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

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

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

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

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

(v) Repeat (iv) for E2 shapes B_{12} through B_{28}.

(vi) Select the MSE mode of gradient calculation.

(vii) Iterate steps (ii) through (v) until the gradient is within ±0.05 dB of zero for all shapes.

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

V. REGULATING REPEATER

5.1 Regulating repeater function

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

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

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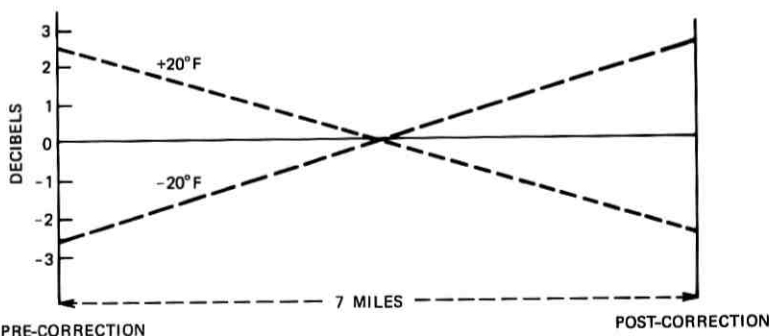

Fig. 21—Regulating section—66-MHz misalignment at the regulating repeater output and input owing to cable temperature variation.

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

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

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

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

Table II — Misalignments allocated to the regulating repeater

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

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

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

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

(i) To equalize the dynamic temperature-dependent term of the cable equation in a pre- and postequalization strategy.

(ii) To accommodate automatically the uncertainties and cumulative spacing deviation in a postequalization strategy.

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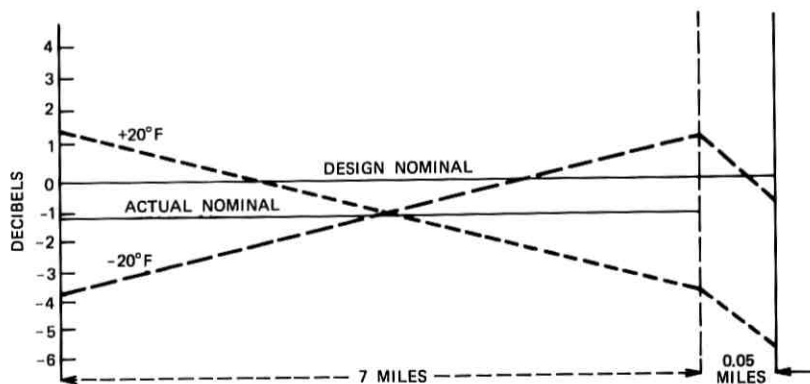

Fig. 22—Regulating section—66-MHz misalignment at the regulating repeater output and input resulting from cable temperature variation and other sources of misalignment.

5.2 Regulating repeater circuits

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

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

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

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

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

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

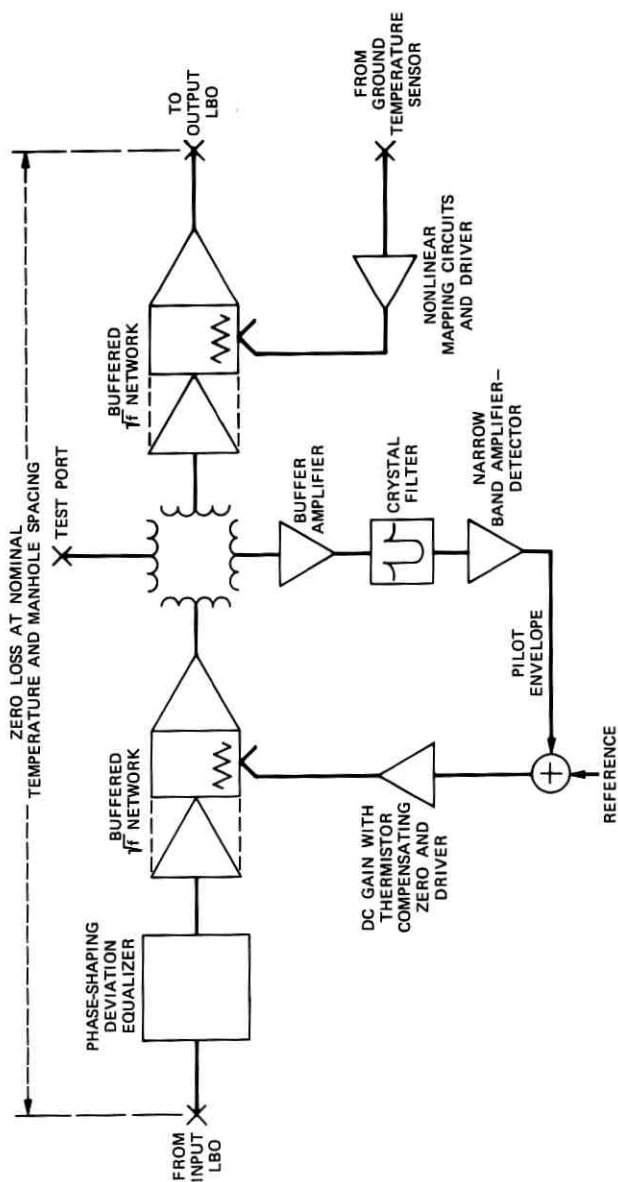(iv) This frequency slot is void of multiplex carrier leaks.

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

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

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

One critical aspect in the design of the preregulator control circuitry is to accurately map the nonlinear resistance-versus-temperature function of the buried ground temperature sensor into a linear loss (in dB) function. In addition, provision had to be made to control the sensitivity of the function since regulating sections can vary significantly in length, and to adjust for the mean annual temperatures since the same sensor is placed in all climatic areas. The response of the temperature sensor is shown in Fig. 24. The range expected is $\pm20°$F
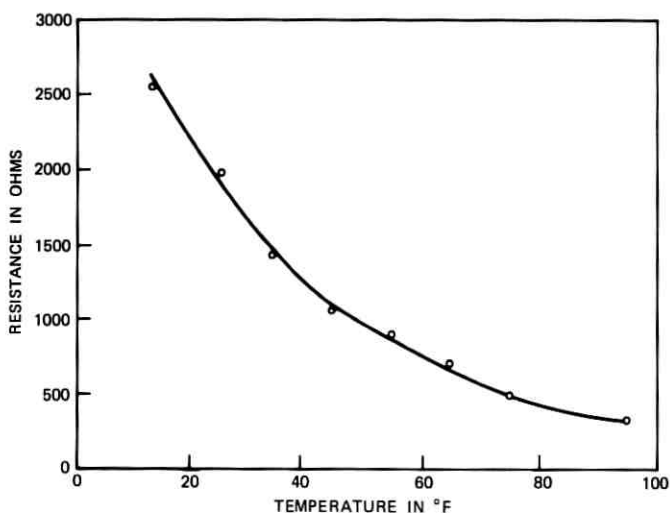

Fig. 24—Ground-sensing thermistor—resistance vs temperature.

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

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

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

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

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

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

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

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

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

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

For very fast variations in the pilot envelope, the envelope gain to the pilot is unity, and the message envelope gain is nearly zero. Message channels, therefore, are not affected by the faster perturbations.
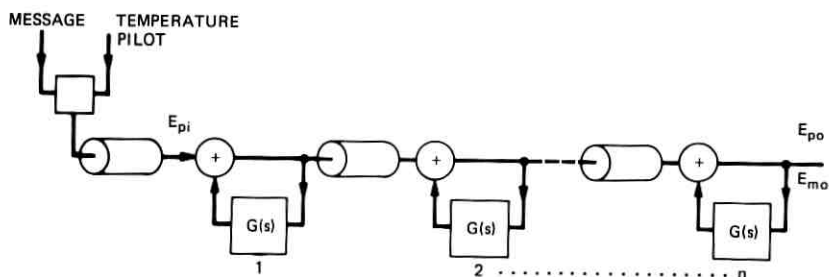
ENVELOPE GAIN OF THE TEMPERATURE PILOT $= \dfrac{E_{po}}{E_{pi}} = \dfrac{1}{[1-G(s)]^n}$

PILOT ENVELOPE TO MESSAGE ENVELOPE TRANSFER FUNCTION $= \dfrac{E_{mo}}{E_{pi}} = \sum_{k=1}^{n} \dfrac{-A \ G(s)}{[1-G(s)]^k} = A \left\{ 1 - \dfrac{1}{[1-G(s)]^n} \right\}$

WHERE $A = \sqrt{\dfrac{f_m}{f_p}}$; f_m = MESSAGE FREQUENCY f_p = PILOT FREQUENCY

Fig. 25—System of tandem postregulators.

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

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

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

One final aspect of the circuitry pertains to a transfer arrangement in the event of pilot failure. The loss of the temperature pilot initiates a switch to the spare line. This pilot loss could also cause the envelope-controlled regulators to go to an extreme gain condition. To prevent this high gain from occurring, heater control of the postregulator is transferred from envelope control to preregulator control. Under these circumstances, those misalignments allocated to the postregulator are no longer equalized. However, in an L5 switching section (150
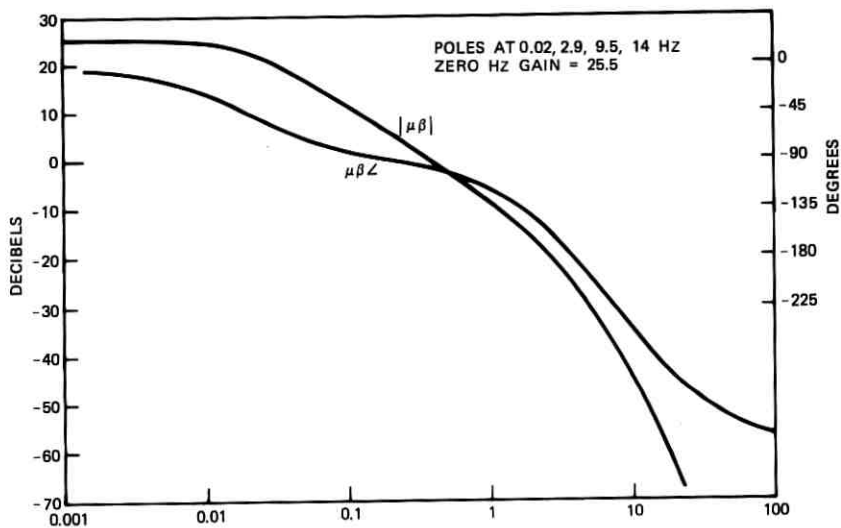
Fig. 26—Postregulator $\mu\beta$ response equivalent to $G(s)$ in the regulator model.
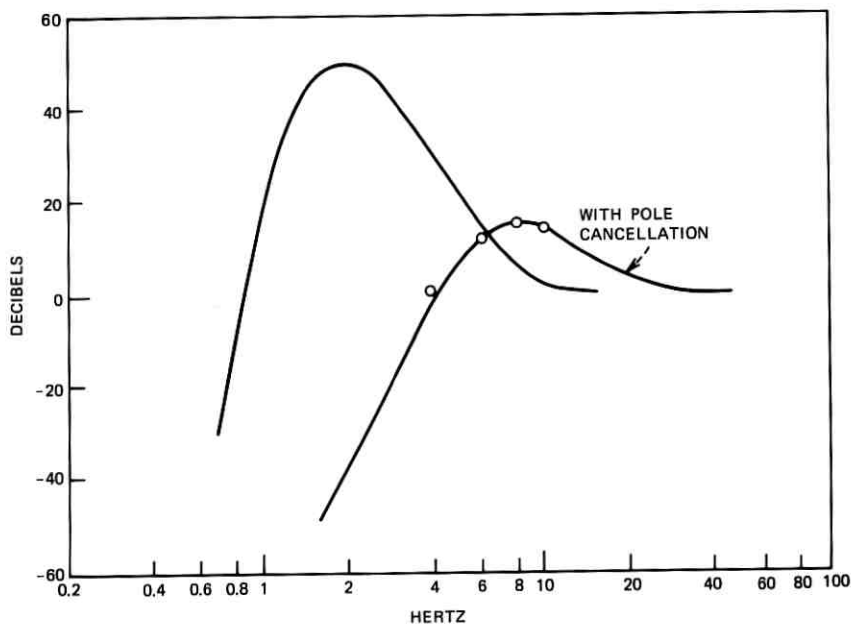

Fig. 27—Computed pilot envelope response for 80 regulators in tandem with and without thermistor pole cancellation. Measured data points are compared.
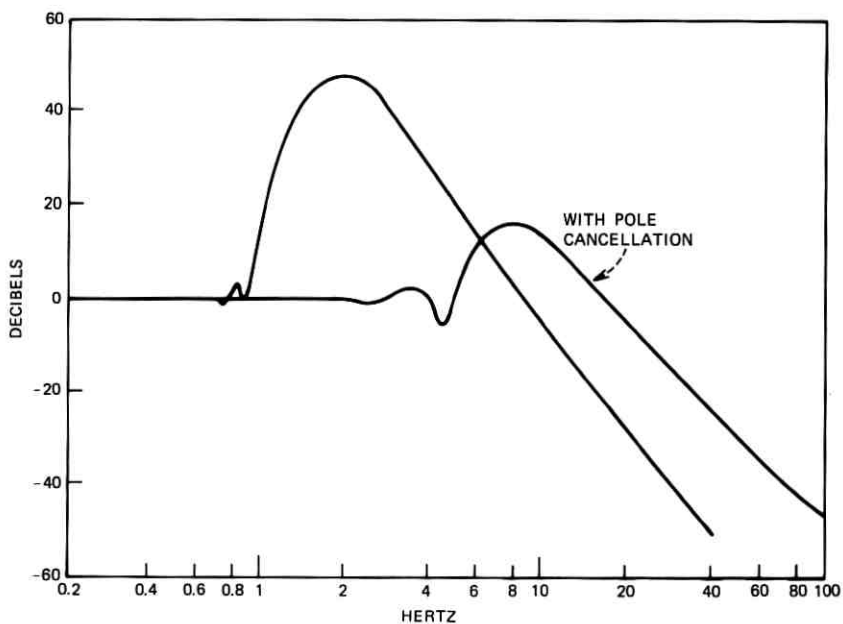
Fig. 28—Computed channel envelope response for 80 regulators in tandem with and without thermistor pole cancellation.
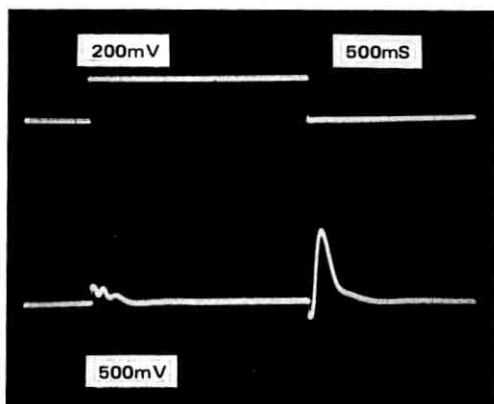

Fig. 29—Transient response of 80 regulators in tandem. Upper trace is the input excitation of 0.8 dB.

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

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

5.3 Regulating repeater noise performance

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

VI. THE E3 DYNAMIC EQUALIZER

The final level of equalization in the L5 system is performed by an automatic pilot controlled equalizer which equalizes for a number of dynamic effects. The primary effect is that associated with the small but significant change in the gain of the L5 line repeaters because of seasonal temperature changes. Repeaters located in manholes are subject to ambient temperature changes roughly equivalent to the $\pm20°F$ temperature change of the coaxial cable itself. Line repeaters having an average temperature coefficient of less than 0.002 dB per degree Fahrenheit at 66 MHz introduce less than 3-dB change in line gain over a 75-mile power-feed section. The variation of this deviation with frequency is predominantly non-$\sqrt{f}$ and is therefore not compensated for by the regulating repeaters.

Table III

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

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

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

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

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

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

All these considerations have led to the E3 dynamic equalizer with four degrees of freedom, each degree of freedom being automatically controlled by a line pilot.
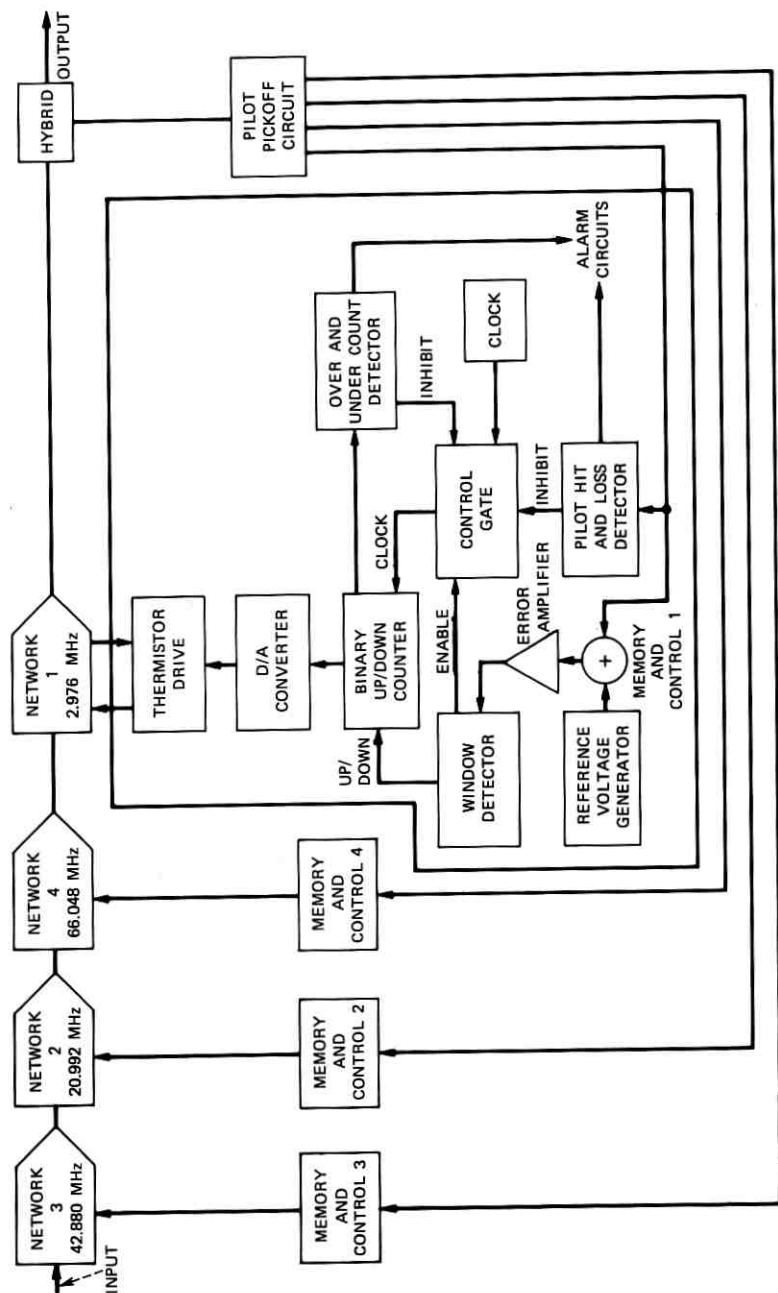
Fig. 30—E3 dynamic equalizer.

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

6.1 E3 equalizer high-frequency networks

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

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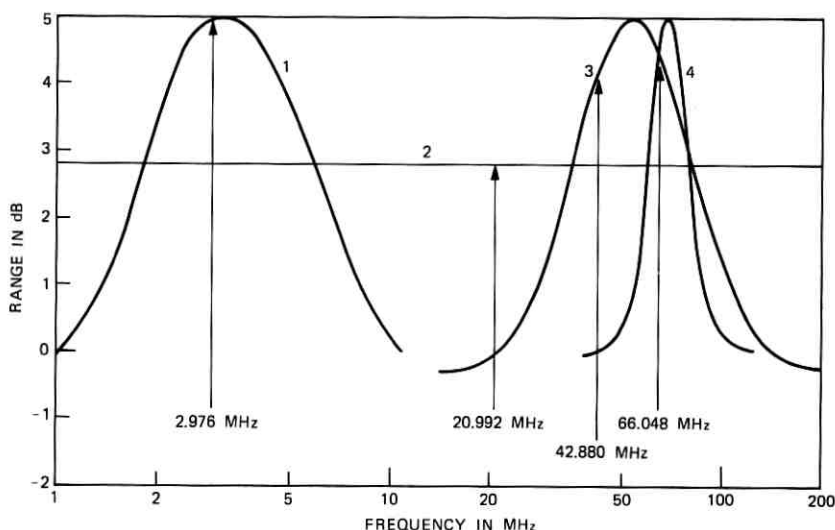

Fig. 31—E3 equalizer shapes.

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

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

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

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

6.2 Control circuits

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

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

    (i) With a digital integrator within the control loop, an equivalently higher loop gain can be obtained than with an analog

integrator. The steady-state accuracy of the control loop is determined primarily by the size of the quantizing steps, the tracking between the digital-to-analog converter, and the network control characteristic (input voltage versus network gain in dB).

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

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

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

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

6.3 Dynamic behavior of the E3 equalizers[13]

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

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

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

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

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

From eq. (12),

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

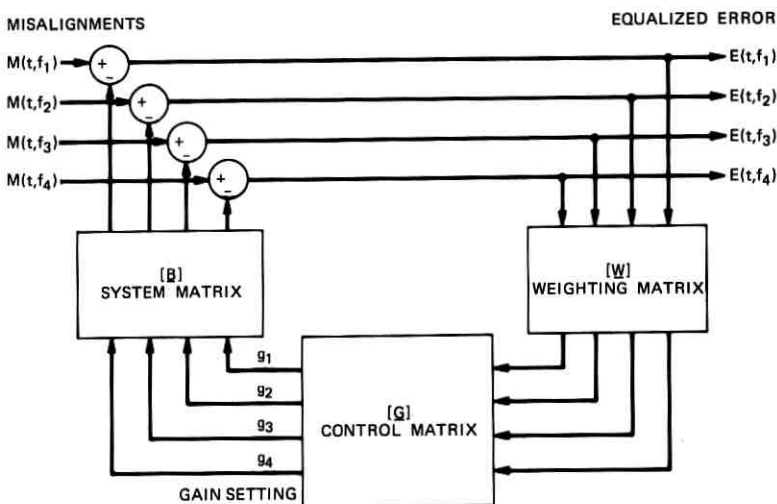where $\mathbf{E} = [E(t, f_1), E(t, f_2), E(t, f_3), E(t, f_4)]^T$, f_1, f_2, f_3, and f_4 are the

Fig. 32—E3 equalizer functional block diagram.

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

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

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

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

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

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

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

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

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

6.4 Response of tandem-connected equalizers

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

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

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

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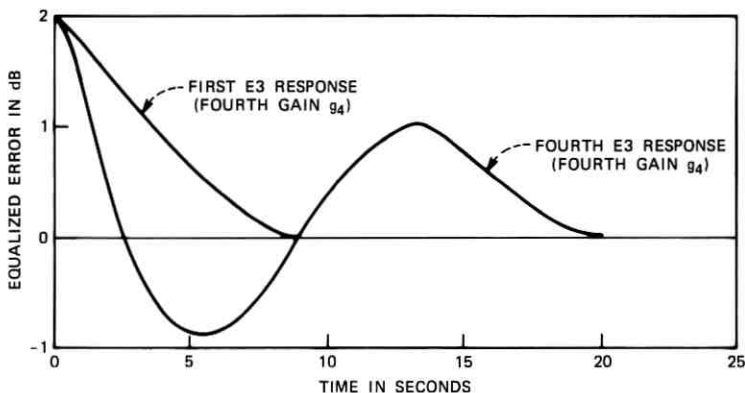

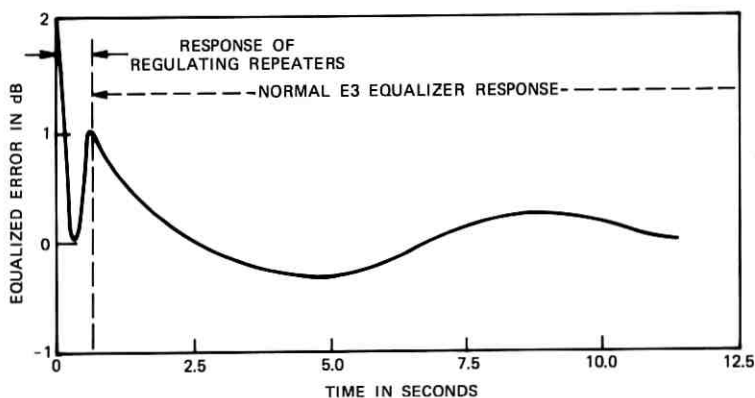Fig. 33—Transient response of four tandem connected E3 equalizers.

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

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

VII. SUMMARY AND CONCLUSION

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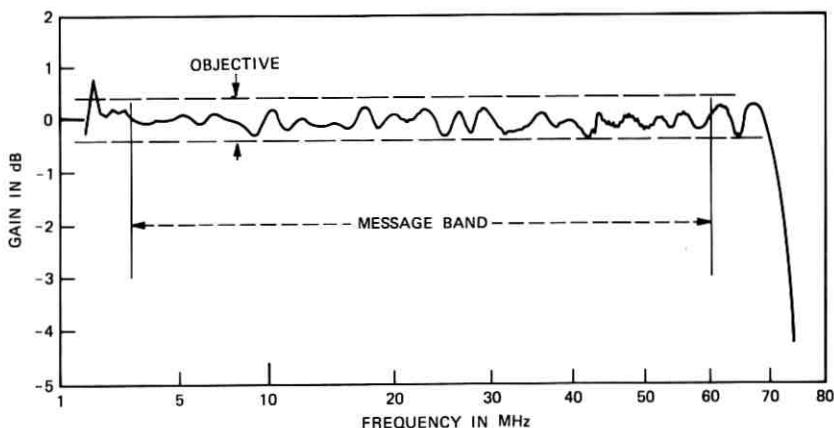

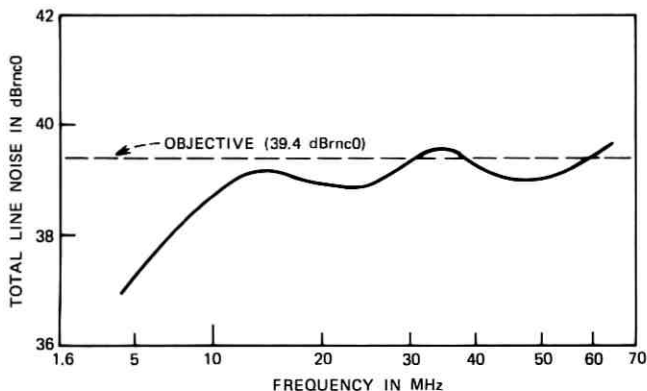Fig. 35—Switching section equalized line response.

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

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

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L5 SYSTEM:

Line-Power Feed

By B. H. HAMILTON and S. MOTTEL

The repeaters in a pair of L5 coaxial-carrier transmission lines are powered in a series loop with two dc-to-dc converters at each end of a power span. The loop is normally grounded at one end with the converter voltages balanced to minimize the voltage to ground at the floating end. As with the L4 system, automatic grounding of the floating point is provided to limit the line voltage to ground and to enable the two lines to be powered independently during turnup and under trouble conditions. A new dc-to-dc converter powered by a 140-V battery has been developed. It employs a single power stage to deliver 910 mA at output levels up to 1150 V. A dc-to-dc converter was also developed for operation on 24-V battery power for use at stations where installation of a new 140-V battery plant is not warranted. The dc-to-dc converters employ hybrid, thin-film, integrated circuits to achieve the needed level of precision without dependence on field adjustments. They also employ many new automatic control features that simplify operation of the four converters in a series loop and permit turnup by remote control via the E2 status reporting and control system. Another new feature is a spare converter with reversible polarity and a decentralized high-voltage patching arrangement.

I. INTRODUCTION

The repeaters and equalizers in the L5 coaxial-carrier transmission system are spaced at approximately one-mile intervals along the cable system. Dc-to-dc converters operating on station battery power and spaced up to 75 miles apart are employed to furnish a constant current of 910 mA at voltages up to 1150 V dc over the coaxial line to the series-connected repeaters and equalizers. This article describes the line-power-feed system and equipment with emphasis on the features that simplify (or automate) system operation and that minimize the

effect of foreign voltages which may be injected into the system between stations.

II. BASIC POWER-FEED ARRANGEMENT

The L5 repeatered lines, like those of the L4 system, are powered in pairs using a series loop including two dc-to-dc converters at each end of a power-feed section (see Fig. 1).[1] For cables shorter than 37.5 miles, the pair of converters at one end is omitted. The system is permanently grounded at one station and the output voltages of the
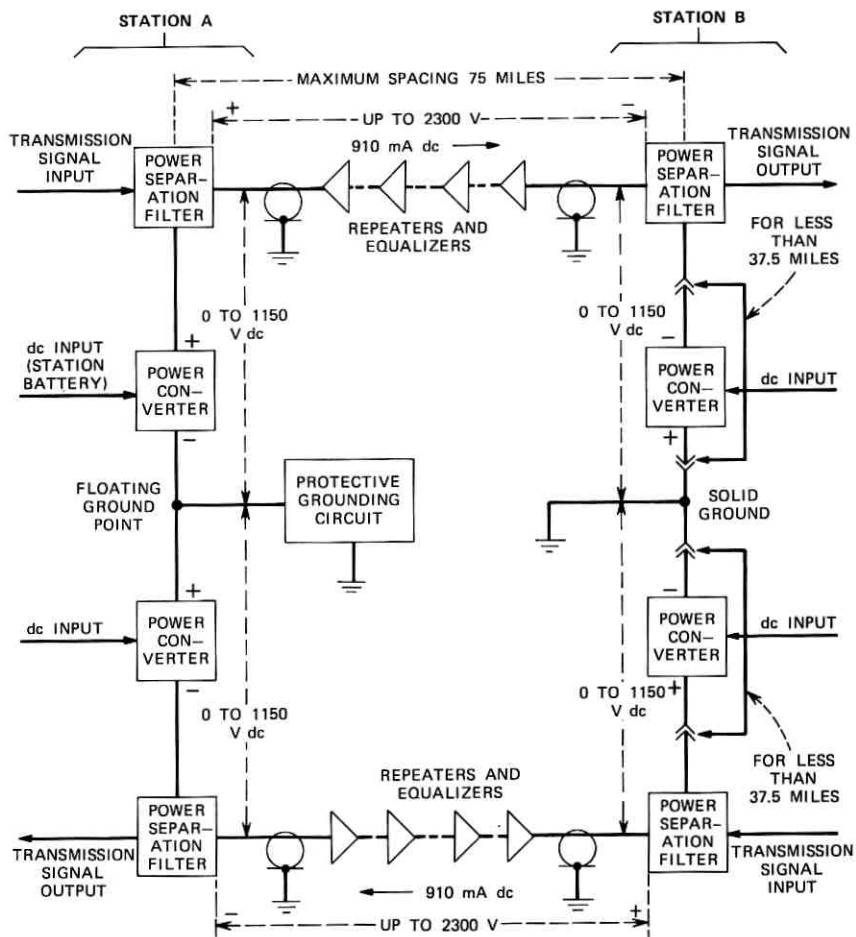

Fig. 1—Power-feed system.

four series-connected converters are balanced under normal conditions such that the voltage is near zero between the floating and local-ground points at the distant station. Therefore, the highest voltage to ground is 1150 V and no path exists for the flow of direct current through earth-metal interfaces between stations. This avoids electrolytic corrosion at such interfaces under normal steady-state conditions.

Under certain abnormal conditions, grounding of the floating point is desired. If, for example, a foreign longitudinal voltage appears between grounded points in the system, grounding of the floating-ground point may be needed to limit the peak voltage appearing between the center conductors and outer conductors of the coaxial cables. This arises from the tendency for such voltages to be superimposed on the output voltage of the converters connected to the floating ground point. A second abnormal condition which warrants grounding of the floating point is the appearance of a cable fault or a converter failure that would reduce the current in the power-feed loop below an adequate level. By detecting such a fault and automatically grounding the floating point, normal current (and transmission) can be maintained in one of the coaxial lines.

III. ELECTRICAL PERFORMANCE REQUIREMENTS

The principal electrical performance requirements that influenced the design of the line-power-feed system are summarized in Table I. The allowance for dc earth potentials as high as 15 V per mile is larger than that used in designing previous systems of this type. Such earth potentials are caused by geomagnetic storms that accompany solar flares. The effects are largest on east-west routes at high latitudes in areas with high soil resistivity. Large dc earth po-

Table I — System requirements

Nominal line-current	910 mA
Loop voltage drop	120 to 4600 V
Line-current variation (normal limit)	±40 mA for combined variations of ±100 V in loop drop, ±2 V per mile dc earth potential, 10–50°C ambient temperature, 120–152 V converter dc input voltage.
Line-current variation (during unusual geomagnetic disturbances)	±110 mA for earth potentials up to ±12 V/mile.
System-withstand capability (no protective shutdowns or equipment damage)	DC earth potential up to ±15 V/mile. 60-Hz induction up to 1500 V rms for 0.1 s.

tentials occur infrequently and persist only for a few minutes in a given locality. The design objectives of avoiding major transmission impairment up to 12 V per mile, and avoiding protective power shutdowns up to 15 V per mile are considered to be quite conservative, inasmuch as earth potentials approaching these levels are expected to occur only rarely and in limited areas of the United States.

High levels of longitudinal 60-Hz induction can be caused by large unbalanced currents under fault conditions on major commercial power lines which parallel the cable route. The "system-withstand" capability of 1500 V is considered adequate for most applications. Where routes have unusual exposures to high levels of 60-Hz induction (i.e., more than 1500 V), additional measures, such as supplementary cable shielding, must be considered.

IV. SYSTEM DESIGN OBJECTIVES

A major factor affecting converter design was the desire for simplicity of system operation. In particular, it was desired to be able to turn up converters at unattended stations by remote control. This implied a need to design the converters and protective ground circuits to perform several functions automatically. These functions include:

(i) Soft start—To increase line current gradually over a period of about 15 seconds.
(ii) Converter voltage limiting—To avoid an excessive output voltage from the first converter turned on.
(iii) Automatic grounding—To ground the floating point during turnup.
(iv) Automatic ungrounding—To unground the floating point upon completion of turnup.
(v) Converter current limiting—To avoid producing excessive current into a short circuit.
(vi) High-impedance shutdown—To prevent turnup into an open circuit (avoids unnecessary personnel hazards).

To provide for faster power restoral after a protective shutdown caused by an unforseeable momentary transient, the converter would be designed to automatically make a single attempt to restart. To maintain the converter and to provide for fast restoral of power in the event of converter failure, a manually patchable spare converter and test-load facility would be available at each power-feed station.

Simple in-service pushbutton tests would be employed to verify that protective converter shutdown and grounding circuits are func-

tioning. The need for adjustments to the system in the field would be minimized.

In summary, it was thought that the goal of high system reliability would be best served by accepting moderate increases in the complexity of the power-feed equipment to simplify its use in the field and permit remote control.

V. POWER CONVERTER DESIGN

5.1 Power conversion

The basic element in the converter is the power conversion section which changes the dc input voltage available from the station batteries to a higher voltage that is controllable and is isolated from the input. Power converters for the L5 system were designed to operate on 24-V and 140-V battery plants. The power conversion technique for 24-V input is similar to that developed for the L4 system,[1] and employs up to five pulse-width-modulated power stages. Each power stage employs power transistors switching at 20 kHz and delivers approximately 200 W at 230 V and 910 mA.

A new type of transistor-converter circuit[2–4] was developed for use with the 140-V battery plants that will be used increasingly in the Bell System.[5] The new converter circuit employs transistors rated at 200 V and 12 A and delivers approximately 1 kW from a single unit. The new circuit, illustrated in Fig. 2, separates the pulse-width-control function from the power-inversion function by employing a switching regulator (Q1, D1) operating at 40 kHz on the input side of a bridge-inverter circuit (Q2, Q3, Q4, Q5) that operates at 20 kHz. A principal feature of this circuit is the presence of the inductor which averages the voltage pulses at the input to the inverter rather than placing the inductor in the conventional position at the output of the rectifier bridge. This provides a number of advantages:

(i) Eliminates voltage and current surges that ordinarily appear in the transistors and diodes as a result of rectifier-diode recovery and/or transformer saturation.

(ii) Simplifies inverter base-drive circuits.

(iii) Reduces number of high-voltage rectifier and filter components.

(iv) Permits effective utilization of techniques to reduce switching loss.

By using nonlinear networks to improve the switching locus of the transistor, the switching losses were kept quite low even at a switching
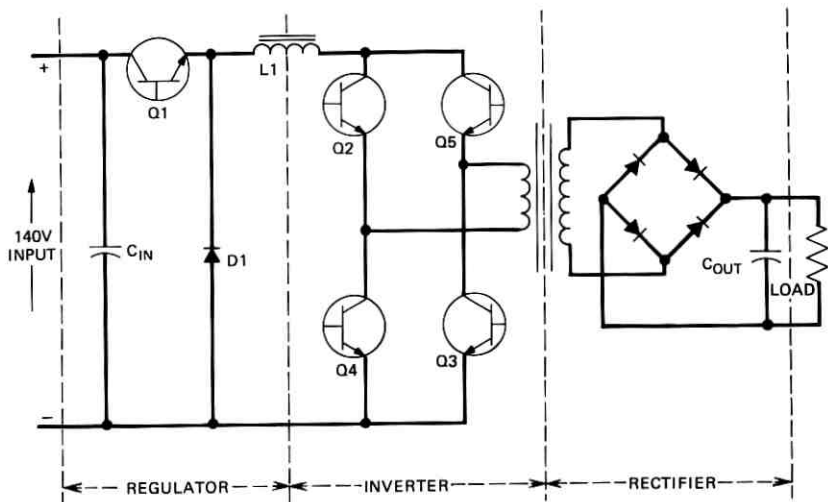
Fig. 2—Power-conversion circuit.

frequency of 40 kHz using transistors with 0.75-μs rise and fall times. These techniques are illustrated for the switching regulator section of the converter in Fig. 3. In Fig. 3a, the switching transistor operates into a highly inductive load (L1) with the inductor voltage clamped by fly-back diode D1. In the absence of the nonlinear loci-control networks, A and B, the adverse switching loci shown in Figs. 3b and 3c are obtained at turn-on and turn-off, respectively, of transistor Q1. The peak and average switching losses in transistor Q1 are 1500 W and 45 W, respectively. Inclusion of networks A and B yields the improvements in switching loci shown in Figs. 3b and 3c. This reduces both the peak and average transistor switching losses by a factor of 10. The total switching losses are reduced by a factor of 3, taking into account the losses in the added resistor, R1. The achieved conversion efficiency of 90 percent (96 percent for the switching regulator alone) is considered quite high at an operating frequency of 40 kHz.

5.2 Feedback control

The desired output voltage-current characteristic for a converter intended to feed a line of maximum length is illustrated in Fig. 4. There are three modes of regulation corresponding to the three linear segments of the voltage-current characteristic.

Normal operation of the converter is in the middle section called the impedance mode. The slope resistance of this section, 3370 ohms,
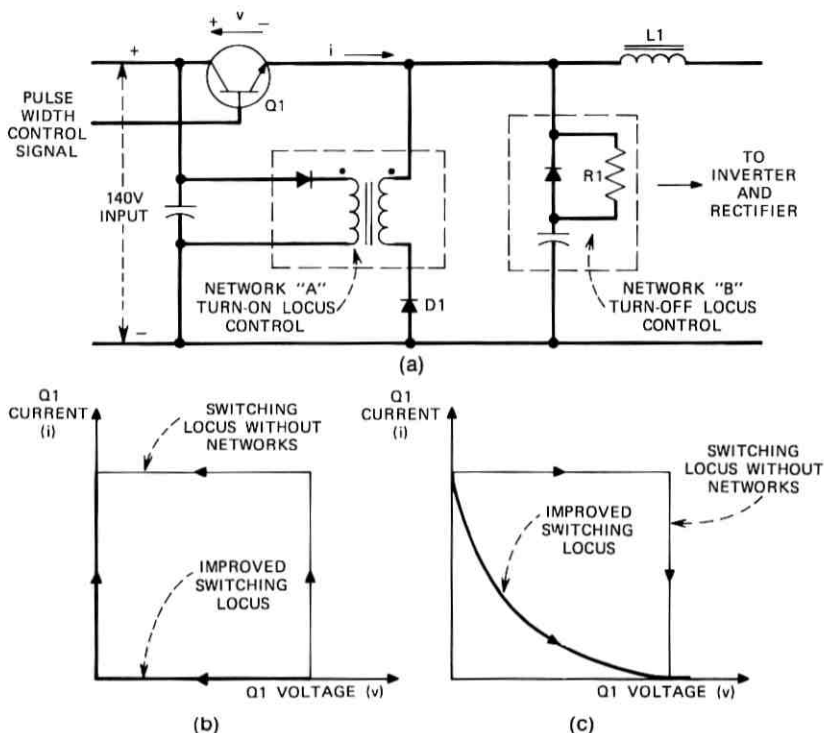
Fig. 3—(a) Switching regulator with locus-control networks. (b) Q1 switching locus at turn-on. (c) Q1 switching locus at turn-off.

is chosen to limit the variations in cable current to ±110 mA in the presence of dc earth potentials up to 12 V per mile. The "level" of the impedance segment is adjustable in the field to accommodate cables of various lengths and to provide means for balancing the voltages produced by the four converters in the series loop. None of the other output characteristics, alarm, or protective shutdown levels require adjustment in the field. Instead, the necessary precision (approximately ±2 percent) is built into the equipment using precise integrated voltage regulators, thin-film tantalum resistors, and integrated operational amplifiers.

At current levels above 1000 mA, the regulating mode automatically changes to the current-limiting segment, where the slope resistance increases to 24,500 ohms. This limits the output current to less than 1050 mA for any short-circuit fault on the line. Normal operation in a
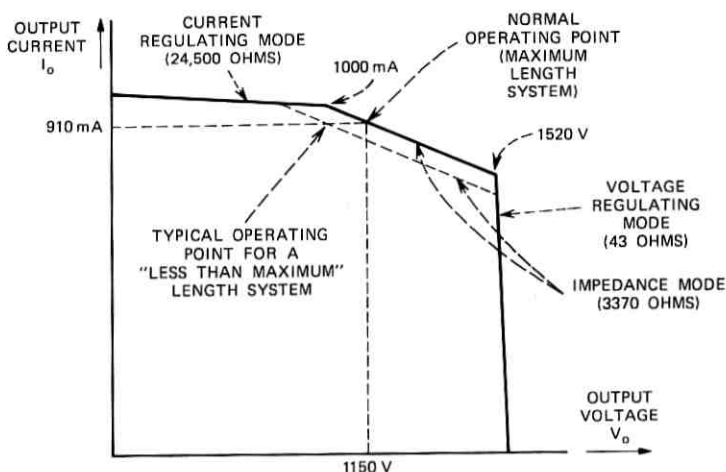
Fig. 4—Converter output voltage-current characteristic.

mode with such a flat (i.e., high-impedance) slope would be undesirable because of the difficulty of maintaining good voltage balancing between the four converters. At output-voltage levels above 1520 V, the regulating mode automatically changes to the voltage-limiting segment, where the slope resistance decreases to 43 ohms. This limits the converter output voltage to 1550 V while power is being turned up. Without this feature, the first converter turned on in the series loop would deliver higher output voltage in attempting to supply power to the complete loop. When the converter is turned on, the output voltage of the converter is increased slowly in the open-loop control mode over a period of 15 seconds to minimize the voltage and current transients applied to the system and to enable the feedback regulator to take control smoothly and restrict the output voltage and current levels not to exceed the characteristic shown in Fig. 4.

The circuit implementation of the triple-mode regulator is indicated in Fig. 5. The output voltage and current of the converters are sensed using saturable reactors to achieve dc isolation. A negative dc voltage proportional to converter output voltage is presented at point a and a negative dc voltage proportional to converter output current is presented at point b. These two voltages are compared to a reference voltage in three different ways at the inputs to each of three "current-summing" operational amplifiers. The outputs of the three operational amplifiers are combined via diodes D1, D2, and D3 to control the

output of a 40-kHz pulse-width modulator which, in turn, controls the output voltage and current of the dc-to-dc converter. The three diodes assure that only one operational amplifier (the one with the highest level) controls the converter. In the current-limiting mode, operational amplifier No. 2 is controlling. Note that the output-current level is compared to a reference level at current summing point c. In the voltage-limiting mode, operational amplifier No. 1 is controlling with the converter output voltage compared to reference at current-summing point d. In the impedance mode, operational amplifier No. 3 is in control. In this case, a linear combination of the output voltage and current signals is compared to reference at summing point e. The slope of the converter output characteristic in the impedance mode is determined by the ratio of the combining resistors R1 and R2.
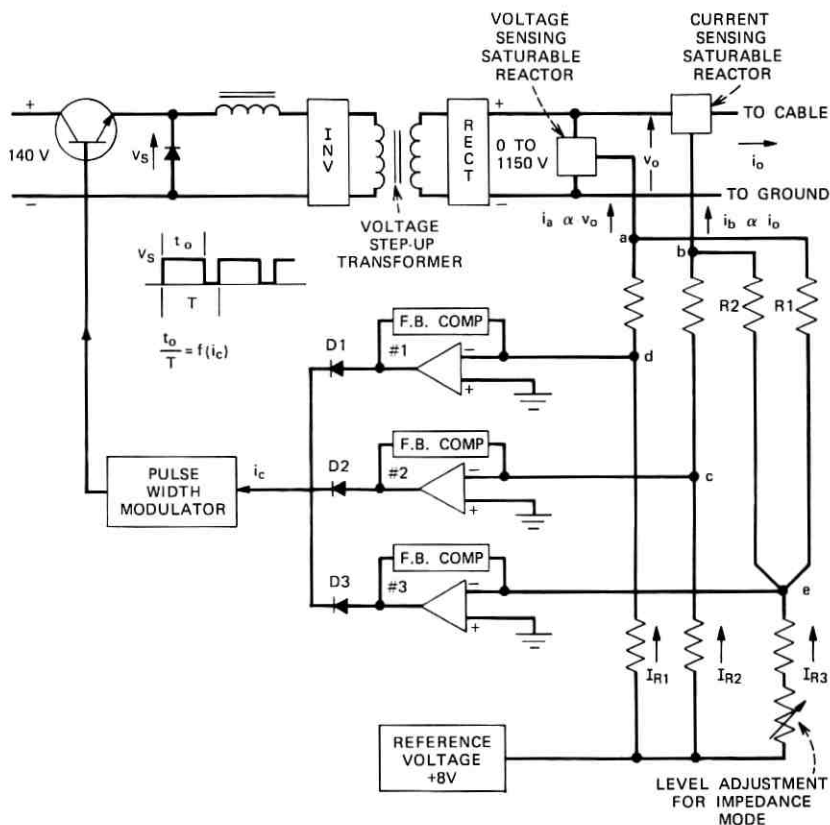

Fig. 5—Triple-mode feedback-regulator circuit.

The feedback regulator circuit and the ramp generator (used for soft starting) were realized as a hybrid integrated circuit using beam-lead operational amplifiers, thin-film tantalum resistors, and ceramic-chip capacitors. The resistors in the current-summing networks are matched to 0.25 percent using an anodizing process.

5.3 Alarm and protection functions

General alarm and protective features were included in the converters as follows:

Low-current alarm (-5%) High-current shutdown $(+32\%)$
High-current alarm $(+7\%)$ High-voltage shutdown (1750 V)

A circuit is also employed to detect line-circuit fluctuations resulting from an arcing fault in the repeatered line and to activate converter shutdown. As a supplementary measure to protect personnel from high-voltage hazards, a circuit is provided which prevents turn-up of the converter if the coaxial line has an open circuit. These functions are realized with three additional hybrid integrated circuits. The precision of these circuits (± 2 percent end of life) is such that no field adjustments are needed. Means are provided for simple (pushbutton) go, no-go in-service tests of these protection and alarm functions.

VI. GROUND PROTECTOR CIRCUIT

The function of the ground-protector circuit shown in Fig. 1 is to automatically ground or unground the system's floating-ground point under appropriate conditions. When the converters are turned down, the floating point is grounded. The ground is automatically opened after two conditions are satisfied: (i) the sum of the magnitudes of the two line currents exceeds 1500 mA and (ii) the difference between the magnitude of the two line currents is less than 34 mA. This assures that all four converters have been energized and that the opening of the ground connection will not produce a voltage at the floating ground greater than 129 V.

The floating point is automatically grounded after a short delay if the voltage exceeds 250 V to ground or instantaneously if the voltage exceeds 800 V ± 15 percent to ground. Possible causes of voltages sufficient to actuate the automatic grounding circuit are a converter failure, a line fault, the appearance of an abnormal dc earth potential between stations, a nearby lightning strike, or abnormal 60-Hz induction. As in the case of other automatic features, a pushbutton,

in-service test is provided to verify the proper functioning of these features.

VII. SYSTEM RESPONSE TO DC EARTH POTENTIALS BETWEEN POWER-FEED STATIONS

As noted earlier in this paper, abnormal solar-flare activity can cause the appearance of earth potentials up to 12 V per mile. Such an earth potential appears as a voltage between station grounds as shown in Fig. 6. The effect of earth potential will be discussed for the case of a maximum-length system where the effect is maximum. Until the earth potential reaches 3.33 V per mile (250 V for 75 miles), the earth potential appears at the floating ground point. The line currents are not affected since the system is grounded at only one point. As the earth potential increases further, the threshold for the automatic grounding circuit is exceeded and the system becomes grounded at both stations. The earth potential now appears as an aiding voltage in one loop (cable plus two converters) and as a bucking voltage in the other loop (see Fig. 6). Thus, in a pair of cables, the cable current increases in one cable and decreases in the other.

Upon taking into account the nonlinear output characteristic of the converters and the major nonlinearities in the repeaters, the overall effects of dc earth potentials are shown in Fig. 7. The hysteresis effect is the result of the margins purposely designed into the automatic
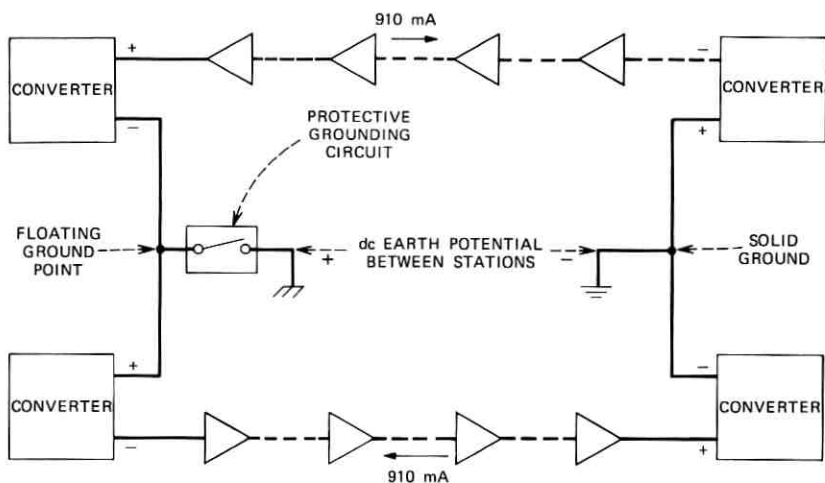

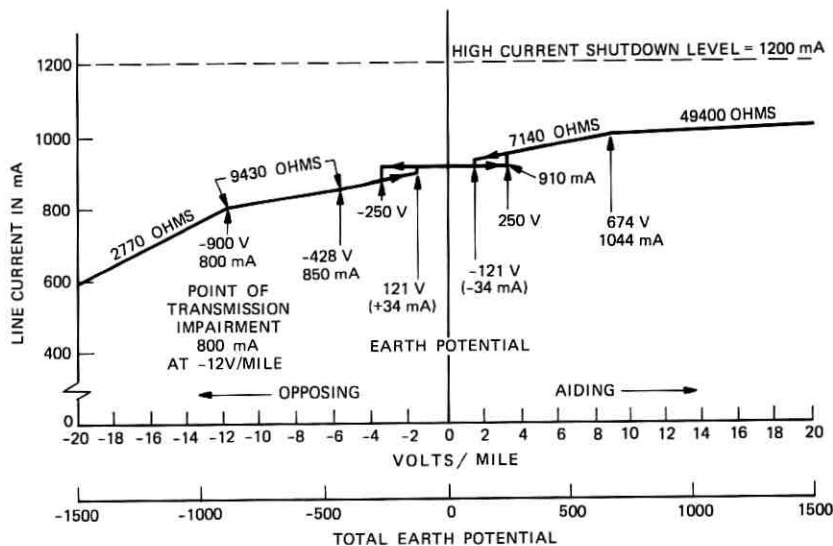Fig. 6—Appearance of dc earth potential between stations.

Fig. 7—Effect of earth potential on L5 cable currents.

grounding and ungrounding circuits to avoid unstable operation. The current variations in the cables are less than 110 mA for earth potentials less than 12 V per mile (900 V in a 75-mile power-feed section).

VIII. SYSTEM RESPONSE TO LONGITUDINAL, 60-Hz INDUCTION

The appearance of a burst of 1500-V, longitudinal, 60-Hz induction near a power-feed station will cause a peak line voltage to ground of 2300 V and a peak line current of 2.2 A. If the 1500-V, 60-Hz induction is closer to the center of a power-feed section, the peak voltage will be less, but the peak current can increase to 3.5 A.

The action of the diodes in the converter output rectifiers is such that the effect on the converter is similar to the application of an alternating series of open and short circuits. Fast current-limiting circuits are included in the converter to reduce the peak currents in the switching transistors. A delay is included in the high-current shutdown circuits to avoid an unwanted power shutdown during such bursts of 60-Hz induction, which are expected to be less than 100 ms in duration.

IX. HIGH-VOLTAGE PATCHING ARRANGEMENT FOR SPARE CONVERTER AND TEST LOAD

A spare converter is provided for use in quickly restoring power to a line in the event of failure of a regular line-feed converter. A manual

high-voltage patching arrangement is provided to permit connecting the spare converter to substitute for the original converter for any coaxial line and, at the same time, connecting the original converter to the test load contained within the spare converter. To achieve this, the output of the spare converter and the test load are multipled to jacks in a protected area at the top of each converter, as shown in simplified form in Fig. 8. There are two types of plugs shown in Fig. 8, P1 and P2. The plugs are shown in their normal positions in Fig. 8. Note that there is a P1 type plug for each converter normally associated with a particular coaxial line, but only one P2 type plug which is normally associated with the spare converter. In the normal state, each individual converter feeds a particular line and the spare converter is connected to its internal test load. Under this condition, the spare converter may be exercised into the test load to assure its operability should it be needed to substitute for an individual converter. The spare converter also provides a means for checking operation of the spare converter plug-in units at each station.

When an individual converter fails, the P1 plug from the failed converter is manually interchanged with the P2 plug from the spare converter. As can be seen from Fig. 8, the failed converter will connect to the test load and the spare converter to the coaxial line. After the failed converter has been repaired and tested, power is turned off on the line and the patch plugs are restored to their normal positions.

The patching is somewhat more complex inasmuch as there are both positive and negative converters for the individual lines and the converter output polarity of the spare converter is reversible. The
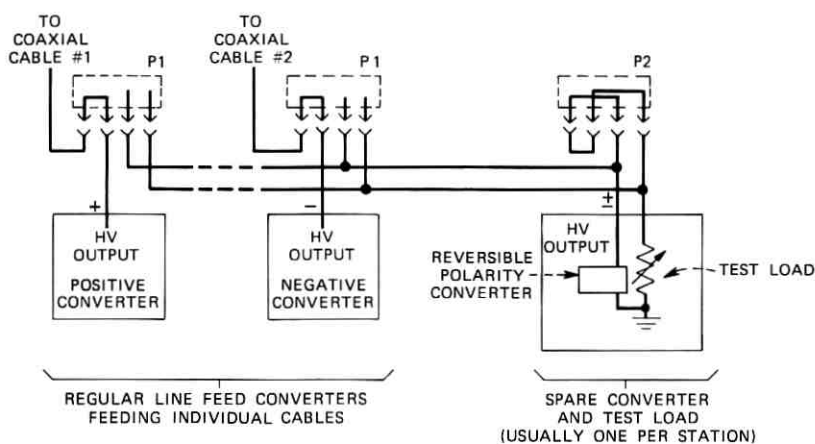

Fig. 8—High-voltage patching arrangement.

P1 plugs from positive converters are slightly different from P1 plugs from negative converters and this difference is used to control the output polarity of the spare converter when it is used to power a coaxial line. There are safety interlock features built into the patch plugs to assure that converters cannot be energized unless the plugs are in valid positions and the covers over the patch panels are closed. Major design objectives for the patching arrangement were to keep it simple and safe for use by the telephone craftsperson in the field and to facilitate the addition of converters without requiring a power shutdown on all cables at a station.

X. PHYSICAL DESIGN

10.1 Design considerations

The physical design of the line-power-feed equipment for the L5 system was based on the following factors:

    (*i*) Two power sources (converters) and their associated ground panel are always provided as a group to power a pair of repeatered lines (see Fig. 1), and thus can be manufactured and subsequently installed in the field as a single entity.

    (*ii*) The two paired power sources (converters) have the same output voltage magnitudes but different polarities.

    (*iii*) The required line-feed voltages can be predetermined and the power-feed equipment can be ordered from the factory with the proper output-voltage capability.

These factors resulted in a physical design which integrates two power sources (converters) and a ground panel into a single structure (see Fig. 9). While the structure has well-partitioned volumes for each function—converters and ground panel—they are not distinct, separable, physical entities.

10.1.1 Specific design considerations

The physical design of the line feed has been influenced, in detail, by other factors. For instance, the L5 line-feed equipment had to be compatible with L4 line-feed equipment in some respects to facilitate station rearrangements when L4 lines were converted to L5. The height and width had to be the same as in L4 or some rational submultiple thereof. The depth of the L5 equipment was designed to be approximately one-half the depth of the L4 power-feed equipment.

BATTERY INPUT
AND COMMON
TERMINALS

HIGH VOLTAGE
PATCH

HIGH VOLTAGE
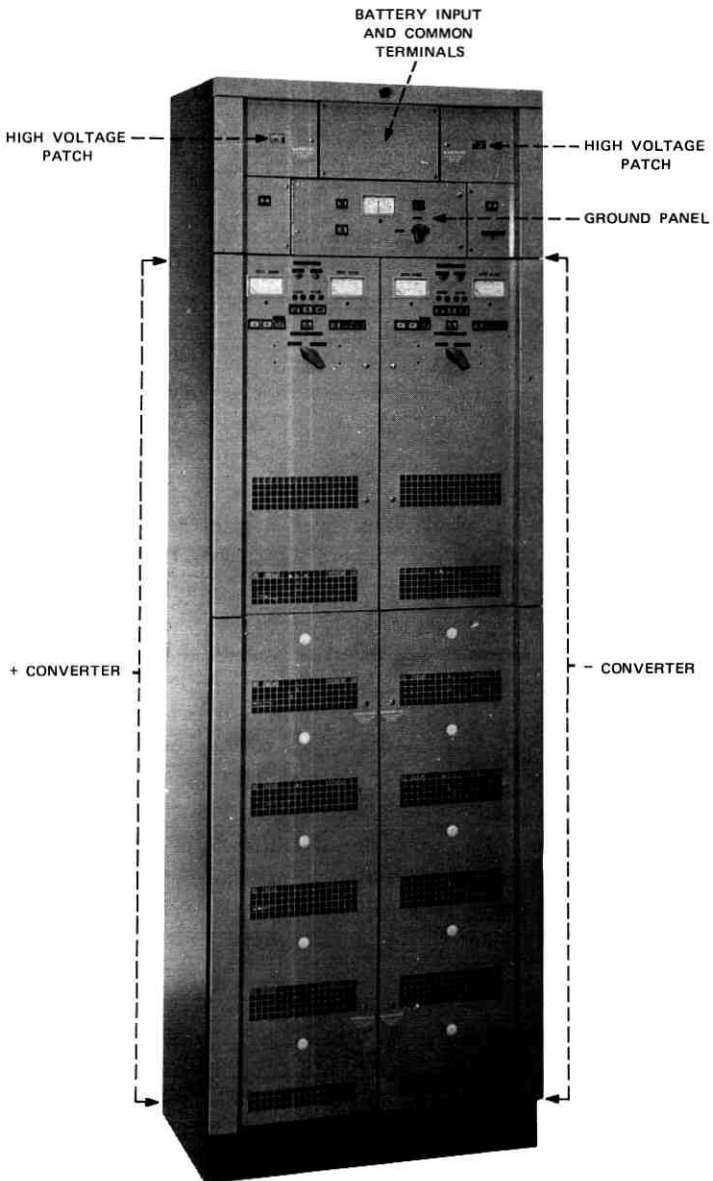PATCH

GROUND PANEL

+ CONVERTER

− CONVERTER

Fig. 9—Line-feed power supply (24-V-battery input).

To facilitate service restoral and maintenance, replaceable plug-in modules and a spare converter were included in the design. Personnel protection was required because of the high-voltage potentials present.

Other design factors were the usual requirements for (*i*) means for heat removal, (*ii*) convenient cabling design, (*iii*) human engineering features, and (*iv*) esthetic considerations.

10.2 Physical design realization

The line-power-feed equipment cabinet shown in Fig. 9 contains two converters and a ground panel. High-voltage patching compartments and a common area for battery input and alarm connections are also provided. Figure 9 shows the 24-V-battery input design, which is 27 inches wide, 13½ inches deep, and 7 feet high. The 140-V-battery input version, designed for stations with the new 140-V dc distribution system shown in Fig. 10, is narrower. The same partitioning pattern has been followed in the 140-V design. Three 140-V cabinets placed side by side are equivalent to two 24-V-input designs and are thus compatible with floor plans.

Normal operating controls are mounted on the front panel. No other facilities for adjustments are provided within the equipment. Trouble indicators are a lamp at the top of the cabinet, and a lamp for each converter.

Figures 11 and 12 show the regular line-feed cabinets with their doors open to illustrate the plug-in modules. The lower doors are interlocked to deactivate the equipment (one converter at a time) since the high voltages are generated in this area. Access doors to other high-voltage areas, including the patching area, are also interlocked and protective covers are furnished.

The cabling plan is designed to minimize cable installation within the cabinet. Field-connected cable terminates at the top of the cabinet. The high-voltage outputs and the connections to the spare converter are protected by flexible conduit and enclosed metal ducts.

10.2.1 Replaceable plug-in modules

Replaceable plug-in modules, which contain the critical electronic circuits, are used to facilitate rapid field repair and manufacture. The modules for a 24-V input converter are shown in Fig. 13. The active electronic circuits for each converter, except for the power stages and oscillator, are contained on five printed-wire boards which provide the plant automatic features described earlier. Each board consists of a double-sided epoxy-glass board assembly with a standard plastic face
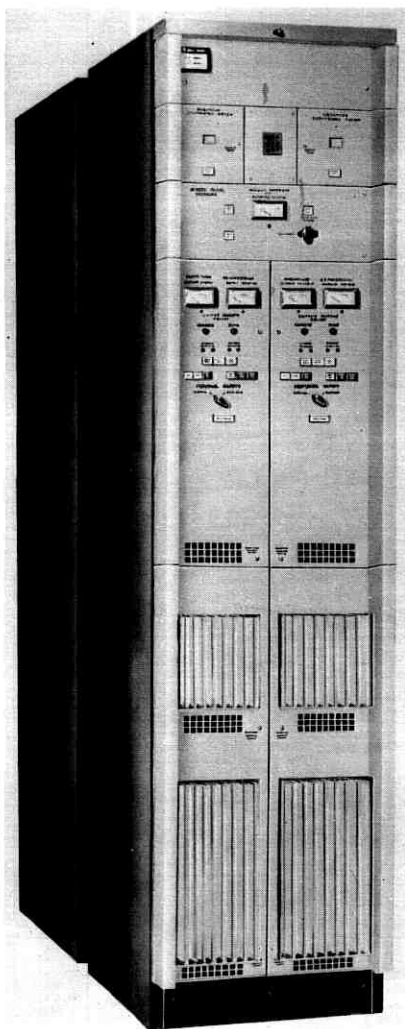
Fig. 10—Line-feed power supply (140-V-battery input).

plate. Connection to the unit is through printed-circuit contact fingers which plug into connectors mounted on the housing for the five boards. Each converter has an identical complement of five boards. Although the boards were initially designed for application in the 24-V-input version, they were applied without change in the 140-V-input version to simplify manufacture and repair. Other printed-wire-

HIGH VOLTAGE
(UNDER PANEL)

HIGH VOLTAGE
(UNDER COVER)

LOW
VOLTAGE

HIGH
VOLTAGE
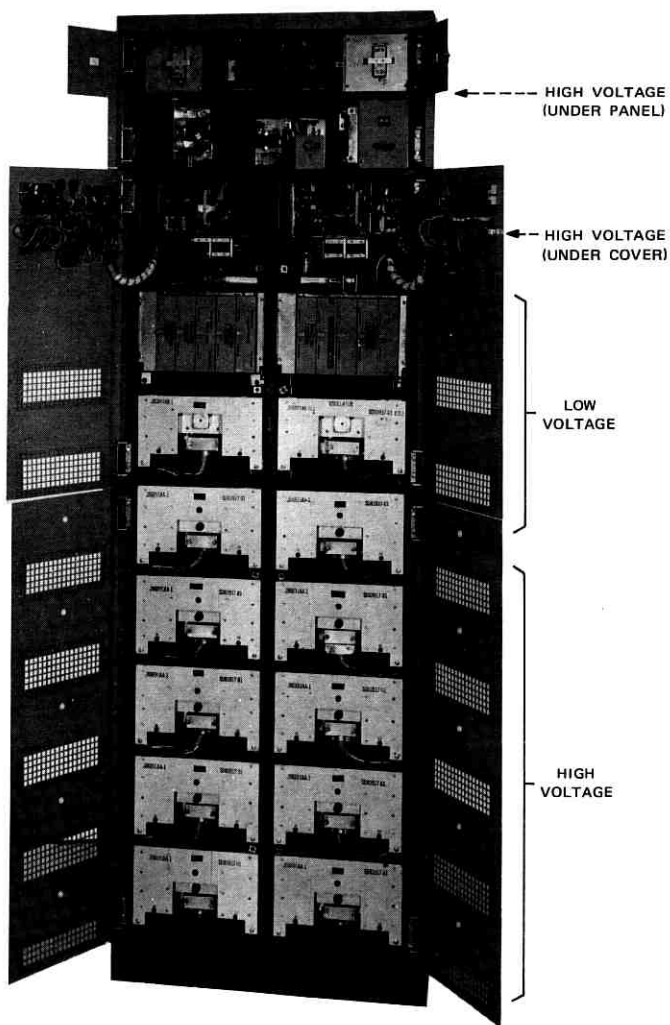
Fig. 11—Interior view of line-feed cabinet (24-V-battery input).

board assemblies are component parts of the power stages, oscillators, and ground panels.

10.2.2 Integrated circuits

The feedback regulation, alarm, and protective features incorporated in the converters utilize four specially developed hybrid integrated
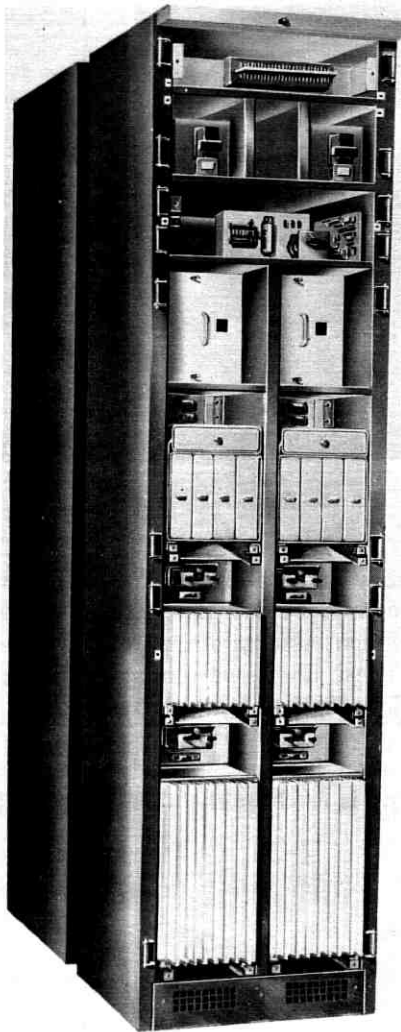
Fig. 12—Interior view of line-feed cabinet (140-V-battery input).

circuits shown in Fig. 14. The hybrid circuits include thin-film tantalum resistors with precise ratios (0.25 percent initial) to provide long-term stability and eliminate the need for adjustments over the anticipated life of the equipment. The resistors, positioned in close proximity on a ceramic substrate, are aged prior to final trim anodization to achieve
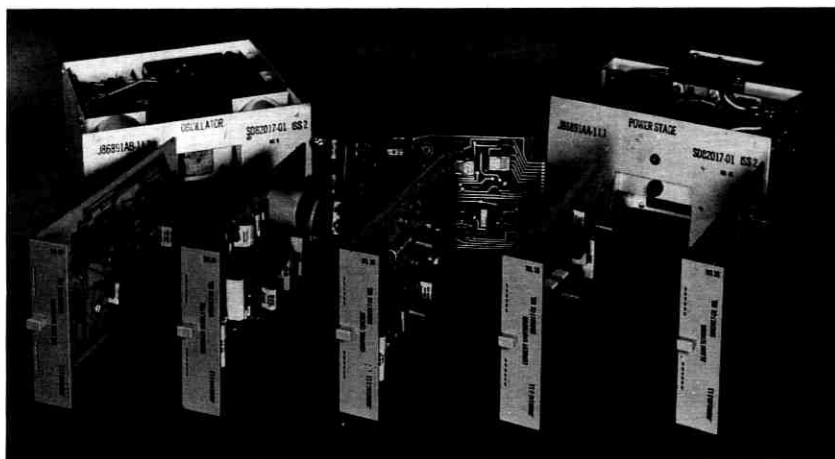
Fig. 13—Replaceable plug-in modules.

the required accuracy and stability. Additional advantages in circuit performance are realized by including associated components on the same substrate, which assures minimum-length paths for critical interconnections. These additional components include beam-leaded operational-amplifier chips, ceramic-chip capacitors, and beam-leaded diodes. Stitch crossovers are also used for interconnections on the substrate. Solder reflow is used to attach the capacitors and thermo-compression bonding is used for the operational amplifiers and diodes
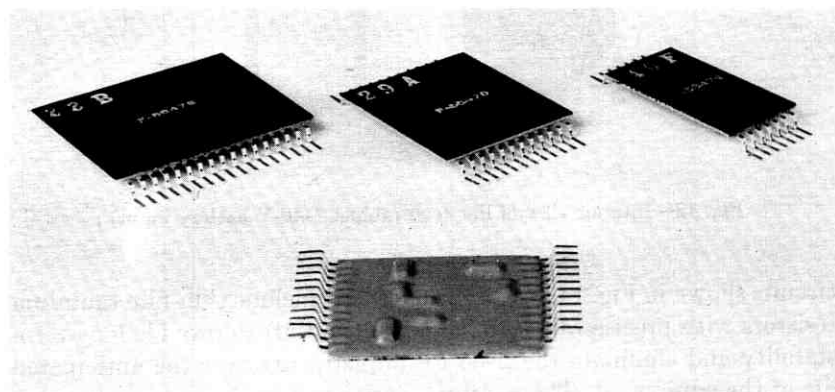
Fig. 14—Hybrid integrated circuits.
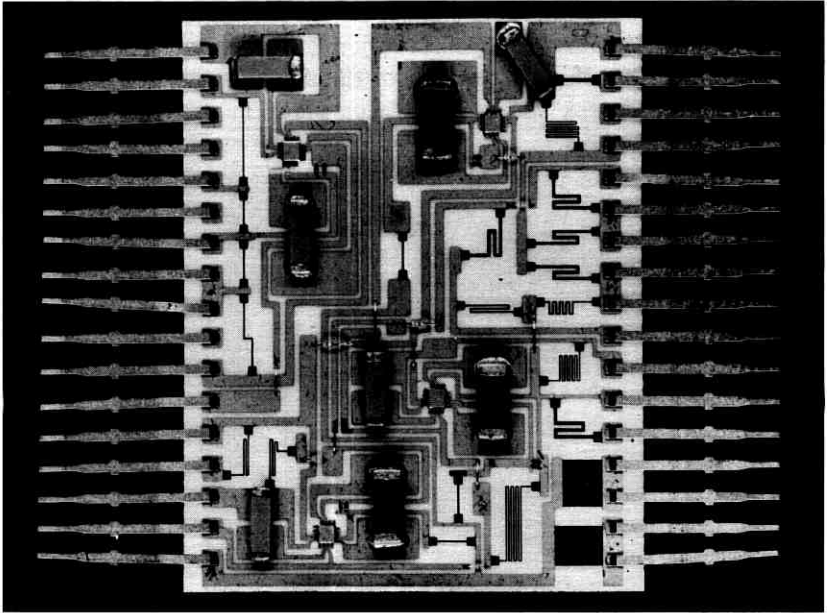
Fig. 15—Unencapsulated development model of feedback regulator with hybrid integrated circuitry.

as well as the substrate lead frame. An early development model of the feedback regulator hybrid, unencapsulated, is shown in Fig. 15. The hybrid circuit is mounted to the printed-wire board by reflow soldering the lead frame to presoldered terminal areas on the boards. The use of hybrid integrated circuitry eliminates a significant number of soldered connections, which would have been required in a discrete realization and thus should enhance reliability.

10.2.3 High-voltage patching and spare converter

The spare converter and the test load used in the high-voltage patching arrangement described earlier are housed in a cabinet similar to a regular line-feed cabinet (see Fig. 16). The spare converter occupies the left half of the cabinet and a resistive test load with its controls, including metering facilities, occupies the right half. Natural convection is used for disposal of heat from the test load.

Each converter (line and spare) has an associated high-voltage patch compartment containing the patch plug and mating receptacle. Access to the patch plug is gained by opening an interlocked hinged

panel, as shown in Fig. 16. A small window in the line-feed converter panel is provided for visual observation of the patch status, i.e., to determine whether the normal plug or the spare plug is in place in the line-feed converter. A status lamp is provided at each line-feed con-

Fig. 16—Spare converter (24-V-battery input) with patch compartment open.

verter and indicates whether the spare converter has been placed in service as a substitute for a regular line feed or is available for assignment.

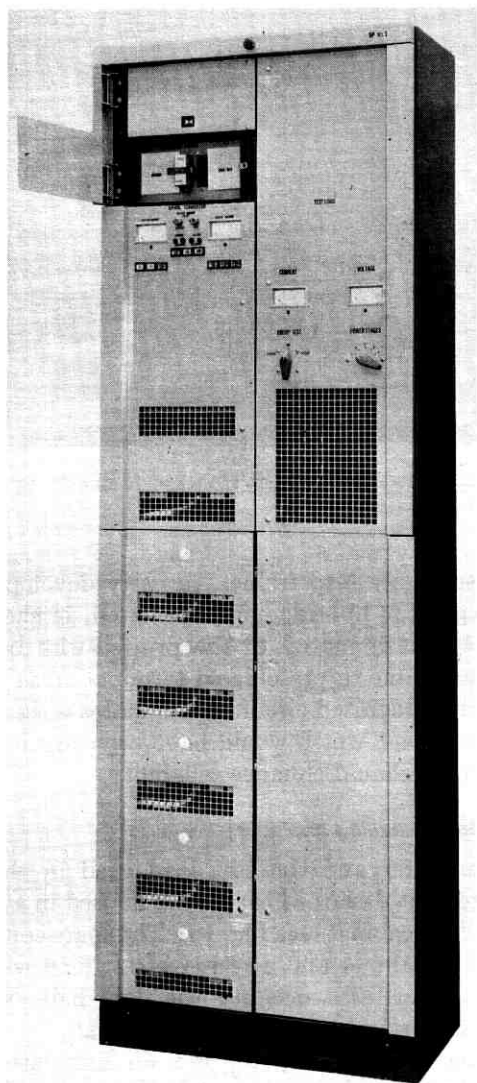The spare converter services a number of line-feed converters, the quantity being a function of the acceptable system repair time. It should be noted here that the transmission-system-protection switching facility maintains service in the event of a converter failure. Arrangements are available to provide a multiplicity of spare converter systems for regular line feed within a single office if system-reliability requirements warrant.

Spare converters are available for -24-V and 140-V input voltages. The cabinet sizes are identical to those of the regular line-feed converters.

XI. SUMMARY

The precision obtained through the use of hybrid-integrated-circuit technology permitted the inclusion of many automatic control features in the power-supply system for the L5 repeaters. It is expected that the resulting simplification of system operation and the minimal number of adjustments required in the field will contribute significantly to overall system reliability.

XII. ACKNOWLEDGMENT

A. P. Walsak was responsible for physical design, and circuit design responsibilities were shared by R. Ostapiak, P. P. Untamo, R. E. Schroeder, and J. R. Meszar.

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L5 SYSTEM:

Line-Protection Switching

By J. H. GREEN and R. W. SANDERS

A new system has been designed to provide protection switching capability for the L5 repeatered line. This line-protection switching system, called LPSS-3, provides one standby line for protection of as many as 10 service carrying lines. Switching equipment is located in terminal, terminal main, and switching power-feed stations. Economic and reliability considerations allow switching spans of up to 150 miles. Either temperature pilot deviations or excessive total signal energy initiates automatic switching. Manual control of all switching functions is provided locally and via telemetry.

A dedicated PCM signaling system maintains interstation switching system coordination. Identical information is transmitted in complementary form over two adjacent L5 channels. Parity and complementary checking as well as automatic retransmission of failed codes enhances error immunity.

I. BACKGROUND

1.1 Motivation for development

The need for new switching capabilities and the desire for a modern switching facility for the L5 Coaxial-Carrier Transmission System have led to the development of a new line-protection switching system called LPSS-3. Several objectives were paramount at the beginning of the new development. One was to design an interstation signaling system that is relatively immune to both line noise and hits (short interruptions of transmission), since a signaling error can cause a service failure because of improper switch activation. Another objective was to provide a design that requires a minimum of effort for the addition of new lines.

A newly installed coaxial-carrier system usually has only one or two of the ultimate regular lines equipped and connected. Additional lines are equipped and connected to the system as traffic growth requires. Thus, the process of adding lines to a working system is important. The LPSS-3 bay is factory-wired and tested for its full capacity of 10 regular lines and one standby line. Equipping switching capability for newly added regular lines on an L5 system requires only replacing dummy plug-in modules with active modules in the LPSS-3 bay and adding a few control leads to the new line bays. No new intra-bay wiring changes or connections are involved.

A new series of coaxial switches has been developed using diodes as the switching elements. These switches were designed specifically to comply with the L5 system bandwidth and modulation performance standards.

1.2 Reliability considerations

A reliability analysis of estimated failure rates, mean time to failure, and service outage was made for the L5 coaxial system. This included estimates relating to the L5 line, main-station, and switching equipment. These studies were then projected to estimate average outage times for various types of switching sections through the use of system models. In addition, the probability of cable damage has been analyzed and is included in the overall system outage predictions. An outage is a service loss, i.e., a failure not remedied by protection switching.

The results of this study show that the use of average switching intervals of 120 miles, total system length of 4000 miles, and one protection coaxial line for 10 regular lines will result in adequately small outage times. With this arrangement, service outage time because of individual line failures is significantly less than outage time caused by massive failures such as a man-made fracture of the buried cable. Modification of switching section length and spare-to-regular ratio will not reduce the latter outage time, which is controlling.

II. GENERAL SYSTEM DESCRIPTION

2.1 Switching section layout

A broadband transmission system such as L5 is composed of two basic elements: the office or main station equipment and the repeatered line equipment. The main station equipment combines and separates the system message content and performs functions such as power feed, equalization, line switching, and fault location. The line equip-

ment provides the transmission equipment required to connect main stations together. The line-protection switching system provides the backup protection to guarantee transmission between main stations when one of the regular lines fails. This is accomplished in L5 by utilizing the LPSS-3 equipment to control and cross-connect the main-station line-connecting equipment to the standby lines provided in the transmission medium. (In the coaxial cable, one unit coaxial is provided for standby or protection use in each direction of transmission, since L5 repeaters are unidirectional.)

Figure 1 is a diagram of an L5 line-switching section. It has been simplified to show only one regular line and the standby line for each direction of transmission. The protection equipment is located within the main stations at each end of the switching section (up to a maximum of 150 miles apart). Within the station, the protection equipment is divided into two basic locations. The line failure detectors and coaxial switches are located in the individual line bays, while the common switching equipment is in the LPSS-3 bay. The inputs to the detector circuits and the access of the interstation signaling commands are derived from the line-connecting circuits.

2.2 Detector circuits

The detector circuits are located in the switch initiator, Fig. 2. These include both transmitting and receiving detectors. The receiving detectors monitor the 42.880-MHz line pilot and the total system average power. If the pilot deviates more than 5 dB from nominal for more than 2.5 ms, a switch request is initiated (whether a switch takes place depends on many conditions and is discussed in detail later). If the average power exceeds its threshold, a switch request is issued, followed 50 ms later by a termination request (as before, the results of these requests depend on other conditions and are covered in detail later).

Transmitting detectors monitor the 42.880-MHz pilot. Each time that pilot exceeds the ±5-dB limit, a 50-ms signal is generated in the LPSS-3 transmitting switch-control circuits, preventing a line switch from taking place during that interval. The assumption is that the failure has occurred in a previous switching section (the probability of such is very high) and that section's protection-switching facility will remove the apparent problem. If both transmit and receive detectors in a switching section persist in the failed state beyond a reasonable switching interval, a switch is completed in that section to obtain the benefit of the independent standby line temperature pilot source.
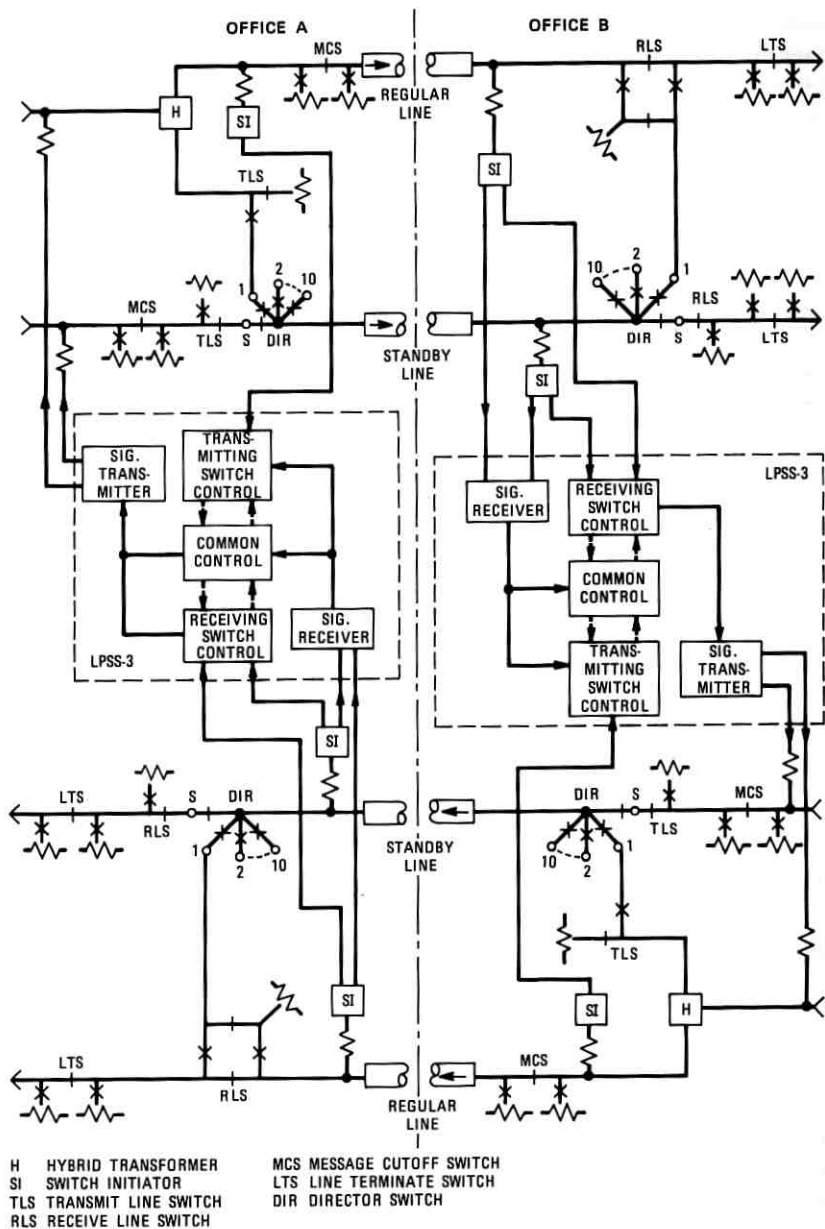
H   HYBRID TRANSFORMER   MCS MESSAGE CUTOFF SWITCH
SI   SWITCH INITIATOR   LTS LINE TERMINATE SWITCH
TLS TRANSMIT LINE SWITCH   DIR DIRECTOR SWITCH
RLS RECEIVE LINE SWITCH

Fig. 1—L5 switching section.
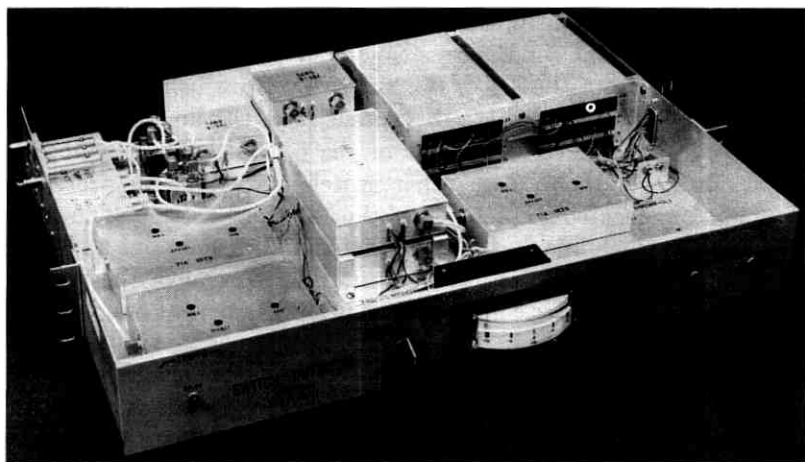
Fig. 2—Switch initiator.

2.3 Coaxial line switches

Two basic types of coaxial switches are used in the protection of L5. One is the director switch, a 1 × 11 solid-state switch without any control of its own; it is a slave to the switches with which it operates. One director is located at each end of the standby line (Fig. 1). Its common port is connected to the line equipment, and its 11 other ports connect to the line-connecting circuits via line switches. With no protection switch in force, the standby line-connecting circuits are connected to the standby line. With a protection switch in force, the directors route the message from the transmitting line-connecting circuit for the failed line, through the standby line, and on to the receiving line-connecting circuit for the regular line.

The remaining line switches are combinations of the basic solid-state T structure. This structure uses two series elements and a shunt element. In the pass condition, the series elements provide low loss and low distortion, while the shunt element has high loss. The stop condition is just the opposite, with the series elements providing high insertion loss to the signal and the shunt element acting as a short to ground for any signal passing the first series element. The line switches are composed of from two to four T's. In Fig. 1, the T's have been simplified to show only their normally released state. (The straight line across the path indicates the pass condition of a T, while the × indicates the stop condition.) The transmitting and receiving line switches connect

the line-connecting circuits to the standby line via the director switches. As pointed out previously, the director switches are slaves to the line switches. In fact, the solid-state elements of the director switches are the final elements of the T's of the line switch to which they connect. Control of the director is by a dc current from the line switch to the director over the center conductor of the coaxial cable carrying the message between them. Two other switches are used in protection switching, one located at each end of the switching section. The transmitting end of each line, both regular and standby, contains a message cutoff switch. The function of this switch is to remove the message load from the line, under controlled conditions, to permit line measurement for special purposes—equalization and, if necessary under high noise conditions, fault location. The line-terminating switch is located at the receiving end of the section, immediately following the receiving line switch. Its function (detailed in Section 4.9) is to prevent the propagation of system overload conditions.

2.4 Circuit devices

High noise immunity and low power dissipation are more important characteristics than speed of operation in LPSS-3 circuitry. With this in mind, saturated logic integrated circuits have been used in LPSS-3 designs. Timing circuits and NAND logic functions are provided by a family of diode-transistor logic (DTL) circuits. Resistor-transistor logic (RTL) circuits are used in the signaling system to provide NOR and EXCLUSIVE OR logic in addition to clocked flip-flops for parallel-to-serial and serial-to-parallel conversion. Set-reset flip-flops from both families are used as memory elements throughout the system.

2.5 Office arrangements

There are two varieties of L5 offices from an LPSS-3 point of view: switching power-feed stations and terminal or terminal main stations. The physical arrangement of equipment in the LPSS-3 bay is the same for either type station, but the manner in which LPSS-3 interfaces with the line equipment differs. The LPSS-3 bay is mounted in the same aisle as its associated line bays. The reasons for choosing this approach are twofold: it was possible to design one universal arrangement that fits all applications, and the maintenance operation is simplified by always keeping functionally associated. equipment in the same aisle alignment. In the terminal and terminal main stations, the LPSS-3 equipment is associated with a line section. In the switching power-feed station, the LPSS-3 equipment is associated with one direction of

transmission in two line sections because the line bays are through-transmission units. No message administration is performed at these stations, so the optimum line arrangement is to have the receiving and transmitting circuits all in the same bay.

III. INTERSTATION SIGNALING

Protection switching operations require coordinated actions at diverse physical locations. Transfer of service from a regular to a standby line, for example, requires properly sequenced operation of line switches separated by up to 150 miles. This interstation coordination is maintained by the LPSS-3 signaling system, a self-contained subsystem that generates and detects digitally encoded carrier signals. The L5 system itself is the medium for transmittal of these signals; the associated frequency band is blocked at each switching main station to eliminate signaling interference among switching sections.

3.1 Encoding and error detection

Thirty-nine digital code words are in active use within the system; each has a specific purpose and consists of seven bits (e.g., 1100101). The digital rate is 2 kilobits per second. When signaling is not taking place, an alternating 1-0 pattern is transmitted to allow immediate alarming whenever interstation signaling continuity is lost. The seven-bit code word is preceded by two successive 1's to mark the beginning of the code word.

The transmitter generates an FSK signal with a 1 bit corresponding to energy at 68.76 MHz and a 0 bit corresponding to energy at 68.78 MHz. The receiver separates and independently detects the two channels. With no transmission or detection errors, the two channel outputs are complementary.

Error-free detection is of prime importance. With this in mind, the 39 code words have been chosen to have even parity to allow detection of errors resulting from the permutation of a single bit. Additional error detection is made possible through complementarity checking of the two channels for each bit of the code word. A transmission medium disturbance must simultaneously permute a 0 to a 1 in one channel and a 1 to a 0 in the other channel in at least two bit positions to cause an erroneous receiver output.

3.2 Automatic retransmission

Error detection at the receiver causes the incoming word to be rejected; no attempt is made at error correction. To prevent switching

system lock-up under this condition, the transmitter automatically retransmits the code word every 15 ms until evidence is received that the command has been properly decoded at the remote location. The sequence is as follows: An *originating* command is transmitted to the remote station to cause the change of state of a line switch. With this action completed, the remote signaling transmitter generates an *answering* command that is routed back to the originating location. The proper decoding of the answering command stops the retransmission of the originating command. At the remote location, cessation of the incoming originating command is taken as evidence that the decoding of the answering command has been successful, and the retransmission of the answering command is stopped. Every signaling operation consists of a round-trip operation as described above to verify the completion of each originating command.

3.3 Signaling system test

Special provision has been made to allow in-service exercising of the signaling system to verify proper operation. Local or remote (via telemetry) manual action causes the transmitter to generate a TEST command, which is decoded at the far end of the switching section. A TEST RECEIVED command is then returned, completing the round trip and causing the signaling to return to idle. Failure of any involved circuitry along the route will cause a TEST FAIL lamp to illuminate at the originating point. This test can be performed at any time, even with a protection switch in effect, and checks most of the circuitry involved with signaling.

3.4 Switch signal routing

Each set of up to 11 coaxial lines in a given transmission direction has an associated signaling transmitter and an associated signaling receiver. At the transmitting end, a switch signal distribution unit controls the application of the switch signals to the transmitting coaxial lines. The signals are normally introduced on to all outgoing regular lines but, as explained in Section IV, the signals are sometimes temporarily introduced on to only the standby line or one regular line. The signaling receiver is electrically connected to only one line by the signaling receiver switch unit, which normally provides connection to the lowest-numbered "good" regular line. In this case, "good" implies the line is in service (e.g., not manually switched out of service) and not failed. Under certain conditions, the receiver is temporarily connected to the standby line.

IV. LPSS-3 OPERATIONS

In this section, we examine the operations of LPSS-3 in detail through the use of flow charts. Line switching, termination, and message cutoff operations are discussed after additional preliminary concepts are established.

4.1 Receiving control

A switching section consists of line switches and associated switching-control circuits at both the transmitting and receiving ends of the coaxial lines. The receiving LPSS-3 circuitry has primary control of all switching actions in that switching section. The transmitting LPSS-3 circuitry merely responds to directions from the receiving circuitry. Switching activity, manual or automatic, is never initiated at the transmitting end.

4.2 Priority

In LPSS-3 jargon, the term, "priority," is associated with internal LPSS-3 operations, which establish the right for an operation to take place. There are two types of priority. A failed regular coaxial line attempts to establish *switching priority* as a first step in executing an automatic switch. Any operation requiring the use of the signaling system must obtain *signaling priority* before being given access to the signaling transmitter. In both cases, priority avoids the system confusion that would otherwise result when independent, nearly simultaneous operations take place. The priority circuits in LPSS-3 (Fig. 3) are N-input, N-output logic circuits designed so that a driven in-
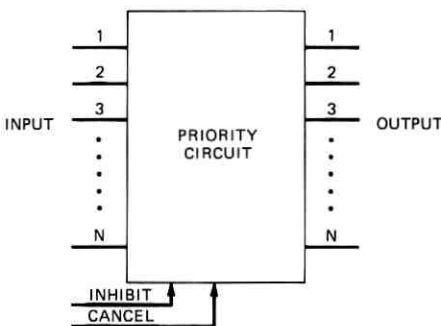

Fig. 3—Priority circuit. Only one output (corresponding to a driven input) may be operated regardless of the input time sequencing. Inhibit prevents new outputs, cancel inhibits new or established outputs.

put will result in the corresponding output going to the operated state. With multiple driven inputs, there will be only one operated output.

If the inhibit input is activated, no new output can become activated; if the cancel input is activated, any previously activated output is cancelled. Consider, for example, the case in which regular line No. 5 fails at approximately the same time as line No. 2. This will result in the No. 2 and No. 5 inputs of the switching priority circuit being driven. Depending on the input sequence, either the No. 2 or the No. 5 output will be driven, but both can never co-exist, even for short intervals. Thus, only one line will obtain clearance to establish a switch. At certain times (e.g., standby line failed) LPSS-3 will refuse to grant new switching priority, but will not defeat established priority. This causes the inhibit to be driven. Both new and established priority may be defeated by the cancel input. Lock normal (Section 4.8) is one such condition.

4.3 Automatic switch

The major steps in accomplishing an automatic transfer of service from a regular line to the standby line are diagrammed in Fig. 4. The line bay switch initiator detects a line failure and automatically causes an LPSS-3 switch request for that line. Switching priority will be established if LPSS-3 can provide switching. Action will stop if another line has switching priority, if the standby line is failed, or if lock normal is in effect. Signaling priority is established next. If another operation is in the process of using the signaling transmitter, another delay is encountered until the transmitter is available. At that time, a specific digital code word called the ID (there are 10 such code words, one for each regular line) is sent to the transmit end of the switching section over all regular lines in that direction. The ID is decoded by the signaling receiver, causing the appropriate transmit switch to operate. An LPSS-3 indication TSO (transmit switch operated) is illuminated. The message is now being introduced to both the regular and standby lines. Signaling priority is sought at the transmit end so that the answering command verifier may be sent back to the receiving end. This one command is used in common by all 10 lines. Receipt of the verifier is taken as evidence that the transmit switch has operated properly. The receive switch is operated, completing the switching sequence. An LPSS-3 indication OSA (out of service automatic) is activated.

The route of the verifier command from transmitting to receiving ends of the switching section deserves special attention. The verifier is
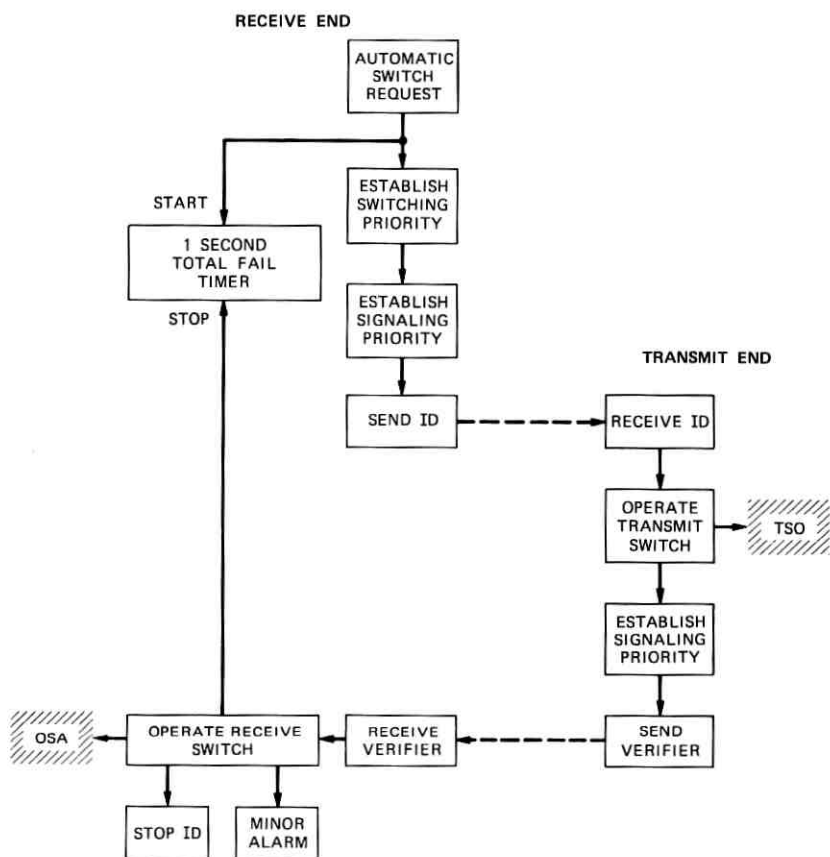
Fig. 4—Automatic line switch.

introduced into the transmit line-connecting circuit of the line to be switched. The verifier then propagates through the operated transmit switch on to the standby line. Meanwhile, at the same time that the ID was sent out from the receiving end, the signaling receiver was connected to the standby line in anticipation of the verifier returning from the transmit end of that line. This routing gives an added measure of confidence in the serviceability of the path to which the message is to be transferred. If the verifier is not successful in traversing this path, the receive switch will not operate.

Elapsed time from failure to switch complete is typically less than 12 milliseconds:

2.5 ms	Recognize pilot failure
2.2 ms	Signaling, ID (zero length system)
0.8 ms	ID propagation (150 miles)
3.0 ms	Delay at transmitting end
2.2 ms	Signaling, verifier (zero length system)
0.8 ms	Verifier propagation (150 miles)
0.3 ms	Delay at receiving end

11.8 ms

Completion of an automatic switch results in a minor office alarm, since switching protection has been automatically used. If, for any reason, a switch request does not result in a completed automatic switch within 1 second, an LPSS-3 TF (total fail) indication and accompanying major office alarm result, since service is presumably lost.

4.4 Switch blocking

In many cases, the 42.880-MHz pilot that is detected for switch initiation traverses more than one switching section. The switch blocking feature prevents several tandem sections from switching on a single failure.

The temperature pilot level in the regular transmitting line-connecting circuits is detected by the switch initiator. When the pilot exceeds limits, the initiator so informs the LPSS-3 transmitting switch control circuitry. The transmitting switch is then inhibited from operation for the next 50 milliseconds. The receiving switch initiator circuits at the next office will also detect the failure, causing an ID to be sent back to the transmit end. Since a block is in effect in the transmit switch control circuitry, the transmit switch will not operate. Under this condition, the BLOCK ON command is sent back to the receive end of the switching section instead of the verifier. At the receive end, the apparently failed line is deprived of switching priority for 100 ms. During this interval, the switching system is free to execute other operations. After the 100-ms interval, the involved line is allowed to seek switching priority if it is still failed. Note that the section with the actual failure is allowed to switch, since its transmitting detector experiences no failure. Subsequent switching sections undergo a 12-ms failure (typical) at both transmitting and receiving detectors and try to switch, but are blocked. After the 100-ms lockout, the lines are no longer failed because of the completed switch in the failed section, and no further action results. If the switch does not complete in the failed section, the next section will switch after the 100-ms interval.

4.5 Automatic switch release

When a failed line that has an automatic switch in force returns to a nonfailed condition as indicated by the switch initiator, an automatic switch release sequence starts. Figure 5 summarizes the operation.

For the first 30 seconds after the line has returned to normal, no action takes place. If the line momentarily fails during this interval, the full 30-second count is restarted. After the line has been nonfailed for the full interval, the receiving switch is released. This transfers service back to the regular line, since the operated transmit switch did not remove the message from the regular line, but merely caused a dual feed of message on the regular and standby lines. The remaining steps are taken to clear the switching system for the next operation. Signaling priority is obtained to send the RELEASE command to the

RECEIVE END

SWITCH REQUEST REMOVED

30 SECOND DELAY

RELEASE RECEIVING SWITCH

START

5 SECOND TSRF TIMER

ESTABLISH SIGNALING PRIORITY

TRANSMIT END

STOP

SEND RELEASE

RECEIVE RELEASE

RELEASE TRANSMIT SWITCH

ESTABLISH SIGNALING PRIORITY

RECEIVE RELEASE VERIFIER

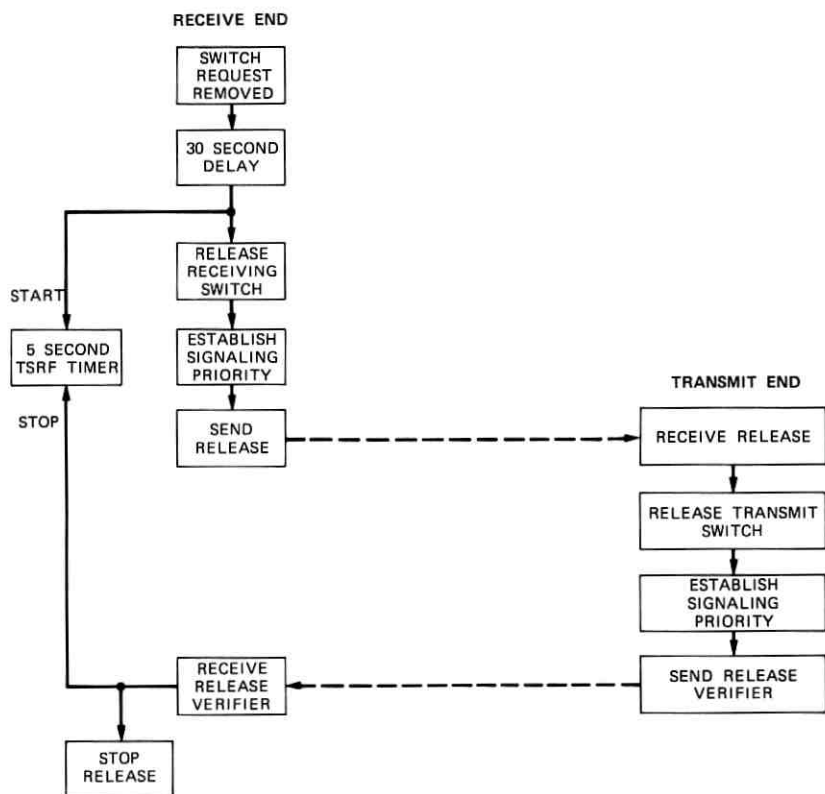SEND RELEASE VERIFIER

STOP RELEASE

Fig. 5—Automatic switch release.

transmit end of the section, causing the transmitting switch to be released. The release verifier is then sent back to the receiving switch control circuitry to verify the successful release of the transmitting switch. If the release verifier fails to return within 5 seconds of the receiving switch being released, the transmit switch release fail (TSRF) indicator is activated, and the minor office alarm is sounded in the receiving office. With this condition in effect, the system will not attempt to establish another protection switch, since a transmitting switch may still be operated. The TSRF condition is cleared by the successful completion of a manual release (Section 4.7).

4.6 Manual switch

Manual switching capability is provided by LPSS-3 so that nonfailed regular L5 lines may be taken out of service for equalization, maintenance, or measurement. The steps taken to establish a manual switch are identical to those summarized in Fig. 4 for automatic switching, except that the action is started manually rather than automatically, the resulting indication is OSM (out of service manual), not OSA, and no office alarms result.

In general, manual operations on LPSS-3 override automatic operations. A manual switch may be executed on one line while another has an automatic switch in force. The manual switch initiation causes the established automatic switch to be released before the manual switch is executed. This capability allows operations personnel to control which of several failed lines is to be switched to the standby line. If the manual switching procedure is executed on a line that has an automatic switch in force, the control of the protection switch is made manual (OSA extinguishes, OSM lights), and the line switch will not release automatically after the failure clears.

4.7 Manual release

The release of a manual switch is identical to the automatic release sequence of Fig. 5 and Section 4.5, except that the action must be started manually and the 30-second delay is bypassed. If a manual switch on a nonfailed line is released with another regular line failed, an automatic switch on the failed line will result. The release of a manually switched line that is failed results in a transfer of control to the automatic mode (OSM extinguishes, OSA lights), but does not cause a release. A manual release may be performed at any time, not only when a manual switch is in effect. This allows operations personnel to clear the TSRF condition, should it occur (Section 4.5).

4.8 Lock normal and restoration lock normal

The lock normal condition is manually initiated to prevent the use of the standby line for service. This feature is used while the standby line is being equalized, for example. The essence of lock normal is that switching is inhibited. Restoration lock normal also prevents switching, but is intended for use when the standby line is being used to carry service from a failed facility that is not normally associated with that standby line. Unlike lock normal, restoration lock normal is activated and released from the restoration patch bay and not from the LPSS-3 bay.

Lock normal may be activated at any time that neither a manual switch nor restoration lock normal is in effect. Restoration lock normal is inhibited by a manual switch, lock normal, or a standby line termination (Section 4.9).

Activation of either lock normal or restoration lock normal results in the release of any automatic line switch that may be in force. Release of lock normal or restoration lock normal allows the failed line to again complete an automatic switch.

4.9 Line termination

The line termination capability of LPSS-3 provides a check against the propagation of overloads on the L5 lines. Each line has a terminate switch located on the output side of the receiving switches (Fig. 1). Each regular line terminate switch may be set in the manual or automatic mode. Under the normal automatic control, any terminate switch that is connected to an overloaded line by the receiving switch matrix is operated, thereby removing the overload from subsequent system components. As previously discussed, an overload on a regular line will cause an attempt to switch that line out of service. The automatic line termination will result only if the switching action (or lack of it) results in a persistent overload condition at the output of the receiving switch matrix. When the manual termination mode is selected, the particular terminate switch involved may be manually operated at any time that the terminate switch is connected to a failed line, where the failure may be due to either pilot deviation or system overload. *Automatic* terminations *result* when switching action cannot stop an *overload*, while *manual* terminations are *enabled* when switching action cannot remedy a *failure*, either pilot or overload.

4.10 Message cutoff

The L5 manual equalization procedure is accomplished out of service, with the message removed from the line facility. A message cutoff

switch is therefore located in each transmit line connect panel. Since untimely operation of this switch while the particular line is in service would cause a service loss, activation of a regular line message cutoff switch is enabled only when the line is manually switched or manually terminated. The normal procedure for removal of the message from a regular line for equalization or measurement is to manually switch the line, then operate the message cutoff control. Enablement of the message cutoff feature for a manual termination is intended to allow the removal of an overload condition from a line to facilitate measurement of that line. Message cutoff capability for the standby line is also provided. Enablement of the feature is caused by lock normal or standby terminate.

Whenever the message cutoff feature is in force, none of the preconditions for message cutoff can be released. For example, if a regular line is manually switched and the message cutoff is activated, the line would be isolated at both transmitting and receiving ends. If the manual switch is inadvertently released, service would be lost. To prevent this, the manual release is inhibited while the message cutoff and manual switch conditions exist on the same regular line.

The steps taken for a regular line message cutoff operation are shown in Fig. 6. Provided one precondition is met when the control is activated, signaling priority is established and the appropriate command is sent to the transmitting end. The command is decoded, causing the operation of the message cutoff switch and the illumination of an indicator MCSO (message cutoff switch operated). Signaling priority is then sought to return the answering signal (message received) to the controlling receiving end of the switching section. When this command is decoded, the MCCO (message cutoff control operate) lamp for the involved line is illuminated, and the release of a manual switch or a manual termination on that line is inhibited.

Release of a message cutoff condition is also executed from the receiving end of the switching section. The message cutoff release operation is never inhibited and may, in fact, be exercised without a message cutoff in effect.

4.11 Pilot resupply

Switching activity often has a direct effect on the presence or absence of pilots on the L5 system. The most critical pilot is the 42.880-MHz temperature pilot. Resupply of the temperature pilot is always provided when that pilot is disrupted by switching, specifically for any regular line message cutoff and for a regular line termination
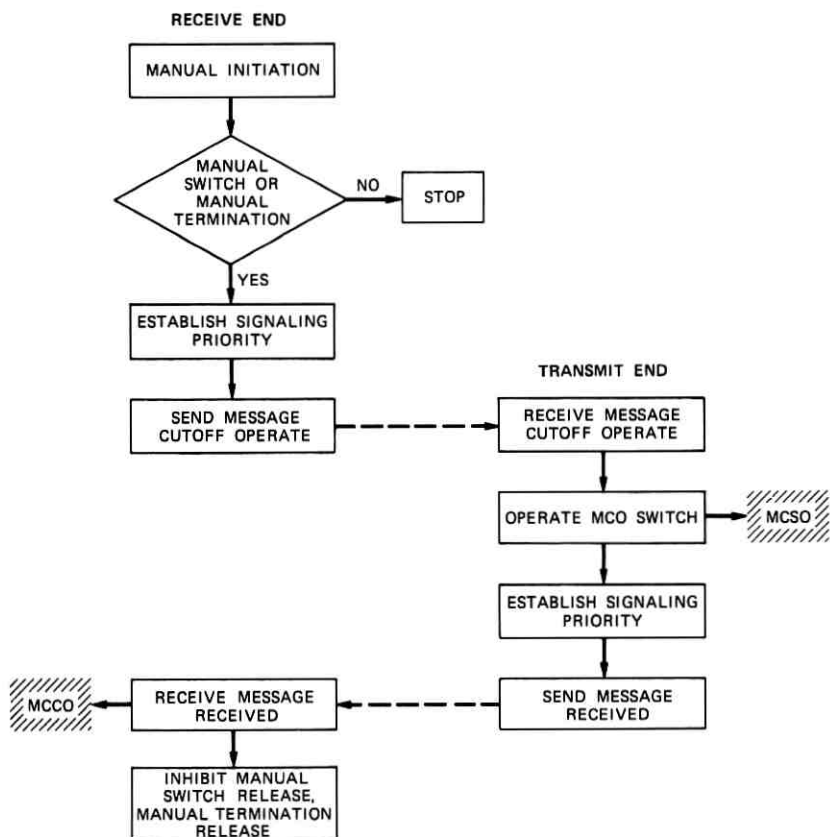
Fig. 6—Message cutoff operation.

at offices where the pilot is not blocked and reinserted in the line-connecting circuit. The temperature pilot is always blocked and reinserted on the standby line on the line side of the message cutoff switch. This eliminates the need for standby line temperature pilot reinsertion. The other three E3 equalization pilots are often passed through the office. Without special provision, these pilots would be lost in the transmitting switching section on the standby line each time a receiving section switch is completed. Because of the relatively high activity of such occurrences, the three E3 pilots on the standby line are reinserted whenever disrupted by receiving switching activity. Except for the 42.880-MHz temperature pilot, E3 pilots are not reinserted on the regular lines, since a line termination is the only means by which

they are disrupted. The combination of the lack of necessity of constant presence of the three E3 pilots and the relative infrequency of line terminations allows this mode of operation.

4.12 Key line feature

One of the 10 regular lines in each direction of transmission may be designated the key line. The key line is different from all others in that a failure on the key line causes the release of an automatic switch on another line, so that the key line may be switched out of service automatically. The key line reacts as do the other lines in all other ways. The key line will not automatically overtake a manual switch, can itself be overtaken by manual action, and will not attempt to switch to a failed standby line. The circuit module with the key line feature is provided only if specifically ordered by the customer. When provided, the capability may be easily disabled by using a switch on the module or transferred from one line to another by simple module interchanging. Figure 7 illustrates the key-line-switching process. If switching priority is available when the key line fails, the line switch to the standby line is executed in the conventional manner. With switching priority unavailable, a release operation will be executed if the standby is not failed, no manual switch or lock normal is in effect, and another line has completed a line switch. When the release is complete, the key line is allowed to establish a line switch. Typical time for a key line switch with another line previously switched is 21 ms from time of failure to key line switch complete.

V. FUNCTIONAL DESCRIPTION

5.1 Subsystem functions

The LPSS-3 bay is illustrated in Fig. 8; one bay is required per coaxial cable end at each office.

Powering equipment is provided in the form of regulated dc-to-dc converters, fuse panels, and power alarm circuits. The converters provide regulated outputs of +25, +12, and +6 volts dc using the −24 volts dc office supply as a source. The various logic circuits throughout the bay have decentralized voltage regulators to provide the proper voltage levels for the integrated circuits.

Signaling circuitry is housed in four shelves; terminal strip and signaling receiver switch unit, switch signal distribution unit, signaling transmitter, and signaling receiver. These components contain all the high-frequency circuitry associated with the generation and distribu-
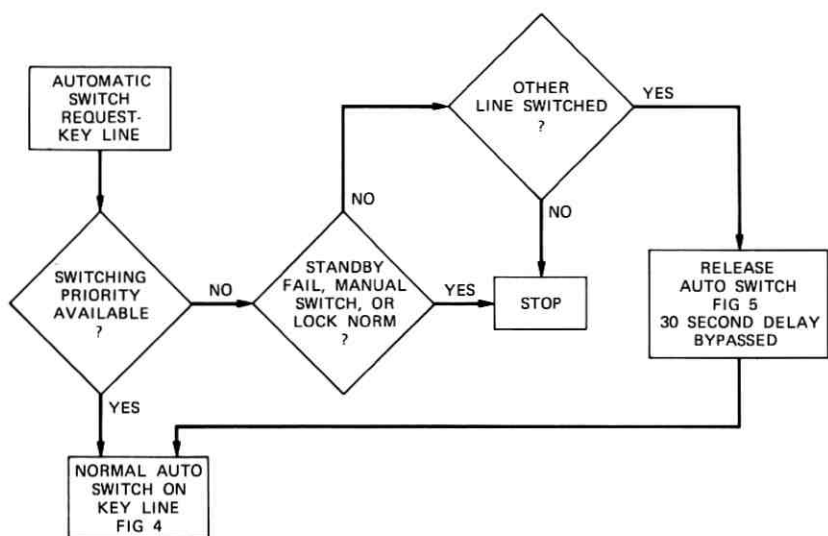
Fig. 7—Key line switch.

tion of the signaling commands and also contain modular logic circuits for the control of signaling.

The *indicator and control* panel provides a centralized input-output capability for routine switching activity. Section 5.2 provides additional discussion of this important panel and associated operations.

Per-line circuits are located on the three shelves directly below the indicator and control panel. The lowest of these three shelves is associated with receiving operations, the middle shelf is associated with transmitting operations, and the top shelf provides interfacing for both receiving and transmitting circuits with the indicator and control panel, the office alarms, and the telemetry systems. Whenever a new regular line pair is added to an existing route, a new working module is added to each of these shelves. Positions for lines not yet equipped are filled by special modules that allow proper switching operation for the equipped lines. Both the transmitting and receiving per-line modules have logic disable circuits with the associated control key and indicator on the module face plate. Activation of this feature prevents LPSS-3 from responding to the normal stimuli that cause automatic line-protection switching. The intended use is for cases in which automatic switching is to be prevented on a per-line basis for switch initiator maintenance or repair and for lines equipped but not yet in service.
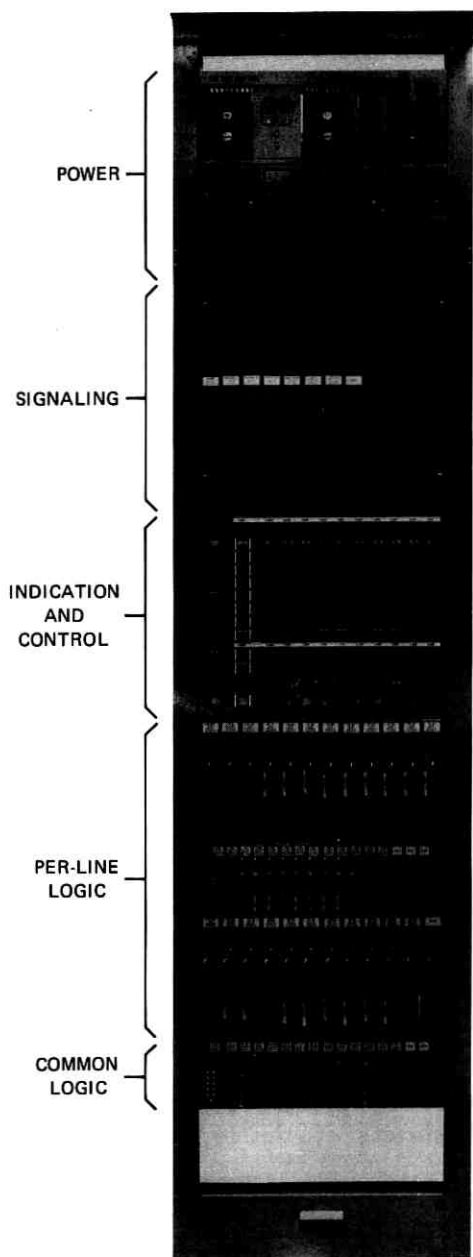
Fig. 8—lpss-3 bay.

Common control circuitry is located in the bottom panel in LPSS-3. The coaxial line to be connected to the signaling receiver is selected by circuitry on the left-hand module. A set of 11 lamps provides visual indication of which line is being accessed. Other functions performed by the common control include switching priority, signaling priority, receive switch timing, release control, standby line receiving control, signaling retransmission, and signaling system test control.

5.2 Indicator and control panel—description

The indicator and control panel (Fig. 9) is the focal point of LPSS-3 manual switching activity and visual indications. The controls and indications are organized in rows and columns by functions. The top eight rows are associated with receiving functions, and the next two rows are associated with transmitting functions. The next-to-last row provides for signaling system control and visual displays. The bottom row is a set of lamp test keys that allow a rapid check of the lamps in the panel.
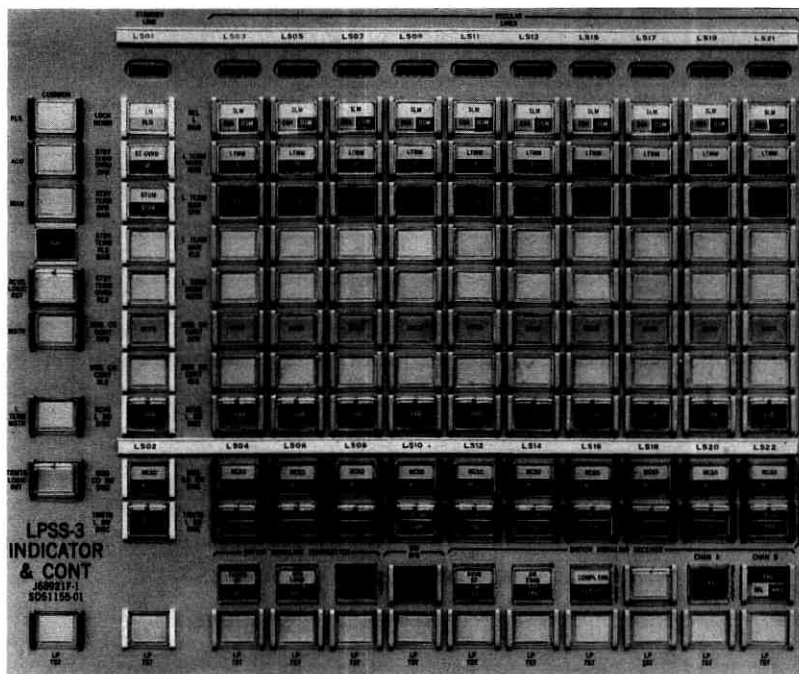

Fig. 9—Indicator and control panel.

The columns contain per-line functions, with the exception of the signaling system row and the left column, which provide common functions. The second column is associated with the standby line, while the next 10 identical columns are associated with the regular lines. Each position on the panel contains a key, a lamp, or a key and a lamp. The key designations are stamped on the panel to the left of the key position, with the exception of LP TST (lamp test), which is below the keys. The designations on the plastic covers are for the lamps under the covers. To illustrate this, the top position in the third column is the SEL L MAN key; the position also contains three distinct indicators, SLM, OSA, and OSM. The upper left position contains the RLS key but no indicator, while the fourth position in the left column contains the TSRF indication but no control key. The control identifications are not repeated for the regular lines. Table I defines the indicator and control panel abbreviations.

5.3 Indicator and control panel—switching functions

The indicators provide office personnel with information regarding the present state of the switching system, while the keys allow the state of line switching, termination, and message cutoff to be manually controlled, as discussed in Section IV. Disconnect keys and corresponding indicators are provided for receiving, transmitting, and message cutoff line switches (e.g., TRMTG L SW DISC and TLSD—third row from the bottom). When a disconnect key is activated, the associated indicator is illuminated and the corresponding line switch is forced to its normal nonoperated state. This allows LPSS-3 maintenance to be carried out without concern for loss of service resulting from improper line switch operation under abnormal switch control bay conditions, such as having circuit modules removed. A message register at the top of each per-line column indicates the number of completed line switches for each regular line and the number of failures for the standby line. These registers are nonresettable, so that the change in readings over a time span is an accurate measure of switching activity.

Two control panel design features reduce the probability of service loss because of accidental manual operations. Particularly sensitive controls, such as the switch disconnect keys, are mechanically inhibited from accidental operation. In addition, manual operations that could interrupt service require simultaneous operation of the involved key and a master key, or the sequential operation of two keys, depending upon the operation.

Table I — Indicator and control panel abbreviations

ACO	alarm cut off
CLK FAIL	clock fail
COM FAIL	command fail
COMPL FAIL	complement fail
DISC	disconnect
DR ENAB	decoder enable
L	line
LN	lock normal
LP TST	lamp test
LSD	line switch disconnect
LTMM	line terminate mode manual
LTO	line terminate operated
MAN	manual
MCCO	message cutoff control operated
MCSD	message cutoff switch disconnect
MCSO	message cutoff switch operated
MSTR	master
OSA	out of service automatic
OSM	out of service manual
OSC A FAIL	oscillator A fail
OVRD	override
PTY FAIL	parity fail
RCVG	receiving
RCVR ST	receiver start
REG LOAD	register load
RLN	restoration lock normal
RLS	release
RST	reset
SEL	select
SF	standby fail
SLM	select line manual
STBY	standby
STOA	standby terminate operate automatic
STOM	standby terminate operate manual
ST OVRD	standby terminate override
TERM	terminate
TF	total fail
TLSD	transmit switch disconnected
TRMTG	transmitting
TRMTR ST	transmitter start
TSO	transmit switch operated
TSRF	transmit switch release fail
WKG	working

5.4 Indicator and control panel—signaling functions

The signaling system control and indication appearances are in the next-to-last row. The first three positions are associated with the signaling transmitter. A normal operation of the transmitter results in a momentary flash of the two diagnostic lamps, TRMTR ST and REG LOAD. An alarm lamp is associated with each oscillator, one for each of the two signaling channels. An alarm lamp is also provided for

the clock signal that times the pulsed high-frequency output of the transmitter.

The fourth position contains the signaling system test key and the TEST FAIL lamp. Depression of the key causes the signaling test operation to start; if the operation has not completed in 1 second, the TEST FAIL lamp is illuminated.

Positions 5, 6, and 7 in the signaling row are associated with the signaling receiver. The alarm lamp COM FAIL indicates that the receiver was unsuccessful in decoding the last command. COMPL FAIL and PTY FAIL indicate a command lacking those properties. Loss of timing ability in the receiver is indicated by CLK FAIL. RCVR ST and DR ENAB are the diagnostics for the receiver that flash momentarily with each normal operation. The last two positions indicate failures of either signaling channel and provide indication and control of which channel is being decoded by the receiver. The channel not being decoded is used for the complement check.

VI. SUMMARY AND CONCLUSION

The LPSS-3 switching system provides protection against service loss because of line failures and protection against overload propagation from any source. Line maintenance is aided through manual switching controls, including service transfer and message cutoff capabilities. One switch control bay provides switching capability for 10 regular lines and one standby line, both transmitting and receiving.

The LPSS-3 bay is modular in design and utilizes a PCM signaling system to maintain interstation switching coordination. Several design features, both electrical and mechanical, minimize the risk of inadvertent service loss from untimely or accidental control activation.

L5 SYSTEM:

Centralized Transmission Surveillance

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A centralized and automated transmission surveillance system has been developed for the L5 Coaxial-Carrier Transmission System as a means for attaining desired transmission reliability. Additional benefits are extensive and include accurate data-processing capabilities plus substantial cost and time economies. The system consists of two basic measuring and control facilities: (i) a transmission surveillance center, located at a designated main station, originates all control operations and accumulates and processes all measured data through use of a small computer, and (ii) transmission surveillance auxiliaries, located at all other main stations, perform measurement functions as directed and return the resulting raw data to the controlling transmission surveillance center. Digitally operated test equipment makes desired measurements under local or remote programmed control, and the E2 Status Reporting and Control System provides interstation transmission of commands and data through time-sharing of a four-wire data transmission link.

I. INTRODUCTION

Verification of the transmission integrity of a complex, broadband network such as the L5 Coaxial-Carrier Transmission System requires many measurements at different locations and considerable processing of the measured data. To minimize the manpower and time requirements for these maintenance operations while providing a high degree of system reliability, a centralized, automated transmission surveillance system has been developed and forms an integral part of the L5 network.

The transmission surveillance system (TSS) takes advantage of a technology evolving from the availability of inexpensive, flexible, small central processors (minicomputers) and programmable test sets. A surveillance network, consisting of a computer-controlled center

located at a strategic L5 main station and a set of remotely controllable auxiliaries located at other stations and having digitally operated test sets with automatic access to key test points, provides the following advantages on a large segment of an L5 route:

(i) Transmission performance overview that is not possible with local individual station maintenance.
(ii) Early warning of slowly developing troubles.
(iii) Rapid localization of trouble by automatic techniques.
(iv) Efficient use of manpower through computer control of routine tasks.
(v) Accurate, flexible records as a result of computer processing and teletypewriter printout.

A four-wire data-transmission system transmits all remote control commands and retrieves remotely measured data in digital form. This data transfer capability is realized by utilizing a new feature included in the E2 Status Reporting and Control System.

The L5 system also has a built-in fault-location capability for remotely identifying a defective repeater in a coaxial line. It consists of an oscillator unit associated with each repeater and a logic unit located in the manhole. Commands from an adjacent main station activate an oscillator unit, and signals injected at the repeater input and output ports are monitored at the receiving end of the line to verify proper repeater operation.

Some additional automated features described in more detail in the following sections include local and remote pilot measurements, line transmission measurements (out of service), computer analysis of measured results, and a system for built-in diagnosis of troubles in the surveillance system itself.

II. OVERALL STRUCTURE

2.1 Functional divisions

The transmission surveillance system (TSS) for the L5 carrier system provides computer-controlled transmission-measuring capability utilizing technology evolving in the field of digitally controlled transmission-measuring equipment. A typical TSS, shown in simplified block schematic form in Fig. 1, serves a segment of several hundred route miles in an L5 system. The main functional units are as follows:

(i) One transmission surveillance center (TSC), the focal point of each surveillance system, originates all automatic operations.
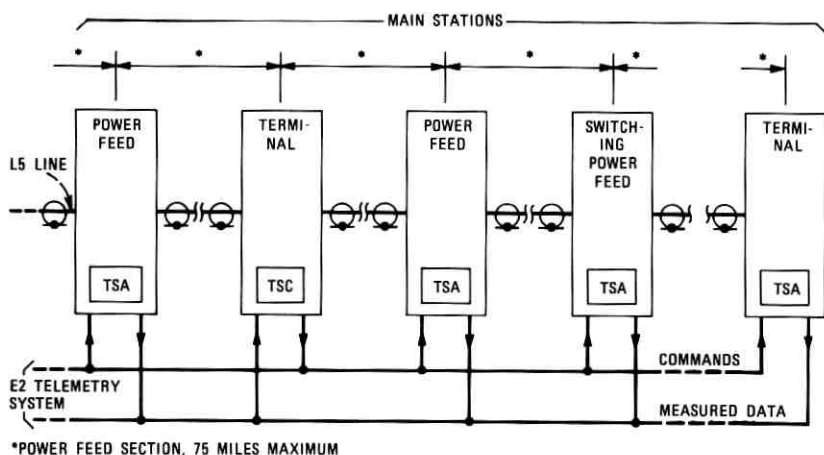
Fig. 1—L5 transmission surveillance system.

*POWER FEED SECTION, 75 MILES MAXIMUM

(ii) Several transmission surveillance auxiliaries (TSA's), preferably not over 10 to 12 per system, are controlled remotely by the TSC and are located at all main stations other than the TSC site.

(iii) A modular, coaxial, switched access network is associated with each TSC and TSA.

(iv) A fault-location facility provides test oscillators at each repeater, remote-control circuits, and power for operating the oscillators.

(v) A data-transfer facility is used for transmission of remote-control commands and measured data, and is provided by the E2 Status Reporting and Control System on a time-shared basis with other services such as alarm surveillance.

2.2 Transmission surveillance center

The TSC is generally located at a well-manned terminal main station having significance in the overall maintenance and operation of the L5 system area covered by the TSS. The TSC functions as a nerve center for originating and processing automatic measurements on L5 line and jumbogroup multiplex (JMX) equipment. Upon diagnosing a trouble condition, TSC personnel may request maintenance action at other distant main stations, which may normally be unmanned or partially
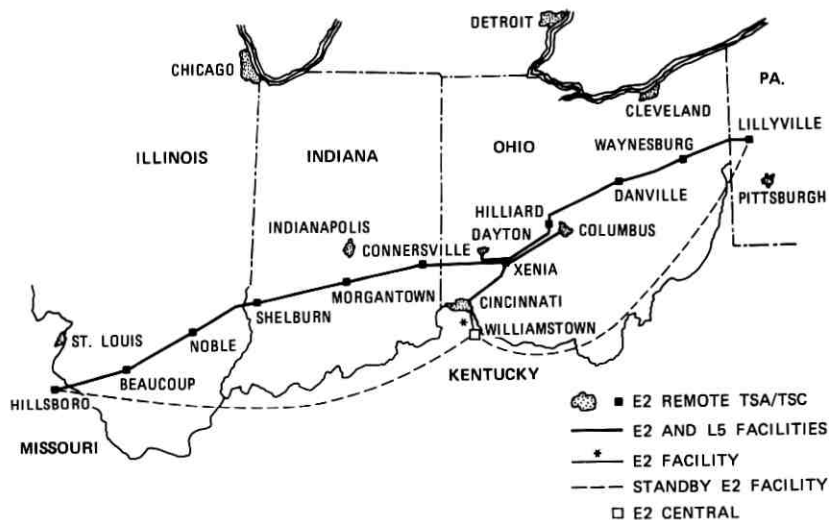
Fig. 2—L5 initial route.

manned. The TSC for the initial L5 carrier installation (Fig. 2) is at the Xenia, Ohio, station, which is on the main backbone route and has three sidelegs and JMX equipment. In this installation, the TSC monitors 13 other stations and 815 miles of repeatered line. Placing the TSC at a station with the largest number of sidelegs and JMX signal-processing equipment maximizes the "local-category" control and measurements. This hastens surveillance operations by minimizing the amount of data-link time sharing required for remote control through the E2 Status Reporting and Control System.

2.3 Transmission surveillance auxiliary

The TSA does essentially the same transmission measurements and local control functions as the TSC, but does not originate automatic commands or process results other than sending raw data back to the TSC. A TSA appears at all station types, which are described in Ref. 1 and listed below:

(i) Power-feed main.
(ii) Switching power-feed main.
(iii) Terminal/terminal main.

2.4 Transmission measuring system

Recently developed 90-type digitally controlled test equipment[2] associated with the TSC and each TSA measures transmission at the

various stations in a transmission surveillance system. The equipment consists of a signal generator providing an adjustable sinewave output in the amplitude range of −99.9 to 0 dBm and frequency range of 10 kHz to 100 MHz. The selective detector measures the amplitudes of received signals in the range of −119 to 0 dBm and 10 kHz to 100 MHz. Most transmission measurements are made automatically under computer control in the surveillance system. A digital control unit (DCU) associated with the 90-type equipment operates the signal generator and selective detector in this mode. Four 16-bit binary words set the output signal amplitude and frequency of a signal generator. Five 16-bit words set the frequency, sensitivity, bandwidth (250 or 2500 Hz), and noise-distortion mode of a selective detector. An analog-to-digital converter in the DCU transforms the analog measurements of a selective detector to digital form suitable for input to the computer or to a data-transmission facility such as the E2 Status Reporting and Control System.

2.5 Command- and data-signal transmission

During automatic surveillance operation, an E2 Status Reporting and Control System associated with the L5 system transmits remote-control commands from a TSC to the various TSA's and raw measured data from the TSA's to the computer in the TSC. The E2 system, as arranged for L5 use, provides interstation communication under the following categories or modes:

(*i*) Alarm polling.
(*ii*) Status reporting.
(*iii*) Remote switching.
(*iv*) Data transfer or remote callup.

Alarm polling, the principal E2 function, continues automatically until interrupted by a request for one of the other functions. Data transfer required for TSS operation is interleaved on a time-shared basis with the other E2 operations. If not inhibited periodically, the remote callup processing of most TSS programs would interrupt the alarm-polling cycle longer than the permissible alarm-updating period. Therefore, the E2 system suspends callup operation every 30 seconds and polls all remote stations for alarms. Callup resumes after a complete alarm-polling cycle, which takes two to four seconds, depending on where the sequence starts.

The E2 system consists of the following units:

(*i*) A central station, which initiates, supervises, and controls most of the E2 system operations.

(*ii*) Remote stations, which may be widely separated.

(*iii*) Remote call-up units (RCU's) (one provided as an integral part of each remote station having general-purpose data-transfer capability).

(*iv*) A four-wire data link for interconnecting the remote stations and central stations.

The manual-type E2 central station provided for the initial L5 installation is located at Williamstown, Ky. (Fig. 2). It accommodates one data link, which may interconnect a maximum of 16 remote stations. At least one remote station is provided at each of the 14 L5 stations for alarm reporting and TSS control purposes.

During alarm polling, the central station automatically interrogates each remote station in succession and registers the station location and category of an alarm should one occur. When an E2-central operator desires detailed status information about a station, he manually initiates the status reporting operations. The frequency and duration of the alarm-polling interruptions for this purpose depend on the number of statuses assigned to each alarm indication and the amount of information needed. Remote switching is used in TSS operations to control the switching of remote line sections for out-of-service measurements.[3] Each remote switch command interrupts the polling cycle for approximately 0.3 second.

During any of the first three E2 system operating modes, the central station and the addressed remote station communicate only with each other. The RCU provides the remote-to-remote communications or data transfer capability needed by a TSC for sending commands to a distant TSA and retrieving remotely measured data. The E2 central station supervises the callup operation, which involves communication among remote stations and between the remote and central stations. The RCU accepts and delivers commands and data in bit-parallel, word-serial form, with each word containing 16 information-bearing bits. As used with the L5 carrier, the E2 system transmits bits serially through the data link at a rate of 600 bits per second.

III. TRANSMISSION SURVEILLANCE CENTER DESCRIPTION

Figure 3 shows the functional arrangement of the principal parts in a TSC. A desk-type console (Fig. 4) contains all the items except two. The standard teletypewriter console is located at the left. The switched access equipment is mounted in the surveillance distribution bay, which is at the left rear in this particular installation.
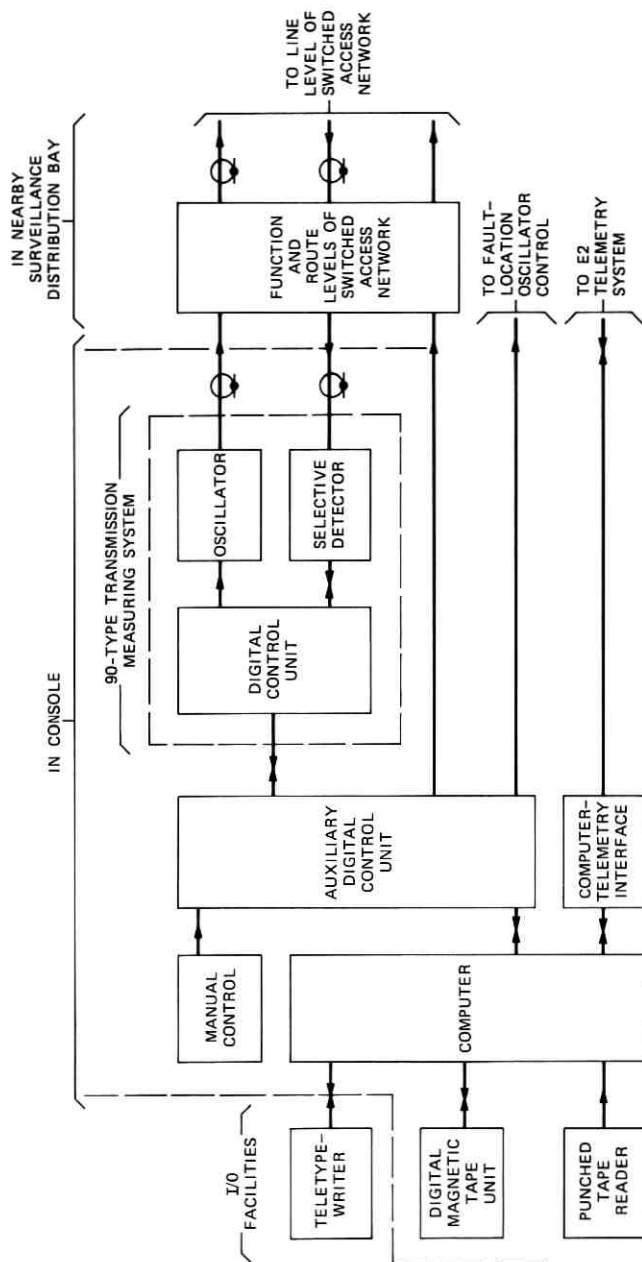
Fig. 3—Transmission surveillance center.

Fig. 4—Transmission surveillance center at Xenia, Ohio, station.

3.1 Computer

The small general-purpose digital computer (minicomputer) in the TSC serves as the central processor in a TSS. The extensive computational and control capabilities in it provide much flexibility for scheduling and sequencing measurements, processing data, and presenting results. Among the salient features of the computer are core memory for 8192 16-bit binary words, a comprehensive program-interrupt system, interface circuits for peripheral equipment, and an internal power supply.

3.2 Teletypewriter

The TSC operator uses the teletypewriter for entering commands and program parameters. The teletypewriter also prints measured results or processed information derived from measured results as directed by the programs.

3.3 Tape reader

The punched-tape reader provides direct computer loading of program material from punched tape. Depending on the type and operational phase of a program, it may be used when initializing the system, when adding information on magnetic tape, when running special programs, or during diagnostic procedures.

3.4 Peripheral storage

Small magnetic tape cassettes, having a maximum storage of 180,000 computer words each, are used in the digital magnetic-tape unit for convenient storage of programs and data. An operator loads tape contents into the computer by typing a command at the teletypewriter.

3.5 Measurement procedure

The digitally programmable transmission-measuring system makes measurements at the TSC location when directed by the computer. The switched access network described in Section V automatically connects the oscillator and/or selective detector to the desired measuring points. Binary-coded instructions pass through the auxiliary digital control unit (AUX DCU) and are held in memory in the DCU to control test set parameters such as frequency, output power, sensitivity, and bandwidth. An analog-to-digital converter in the DCU encodes the measurements into digital form suitable for input to the computer.

Logic in the AUX DCU transfers and steers control commands from the computer or manual control circuit to test equipment, fault location, or switched access control circuits. Flip-flop memory holds the fault location or switched access commands until the command selection is changed or released.

3.6 Computer backup

Front-panel pushbutton keys and associated programmed logic in the manual control provide manual backup of local switched-access and fault-location oscillator control. Numeric readouts display the selections made with these keys. Other keys permit generation of any 16-bit binary word for testing or limited operation of the transmission measuring system. Presently, manual control is possible only on the local-office operated equipment. Subsequently, remote manual control capability will be provided for operating fault-location control logic at adjacent stations and one beyond the adjacent stations.

3.7 Computer-E2 system interfacing

The computer-telemetry interface circuit and a software program driver adapt one 16-bit computer input/output (I/O) channel to the control format of the E2 Status Reporting and Control System. The time multiplexing in the interface circuit and software programming enable the computer channel to receive 16-bit information-bearing words from the E2 system and, in addition, to monitor six E2 system control leads. The interface circuit transfers outgoing 16-bit words from the computer without modification, but does stretch the outgoing strobe pulse (device command) to the duration required by the E2 system.

IV. TRANSMISSION SURVEILLANCE AUXILIARY DESCRIPTION

The TSA has all the functional units of a TSC except the computer and associated I/O peripherals (Fig. 5). A bay framework mounts all equipment except the transmission-measuring system, which is assembled in a rolling console and plugged into the bay. The console may be disconnected temporarily from the TSA and used for general-purpose measurements anywhere in the office during visits by maintenance personnel. The TSA bay also contains portions of the switched access network.
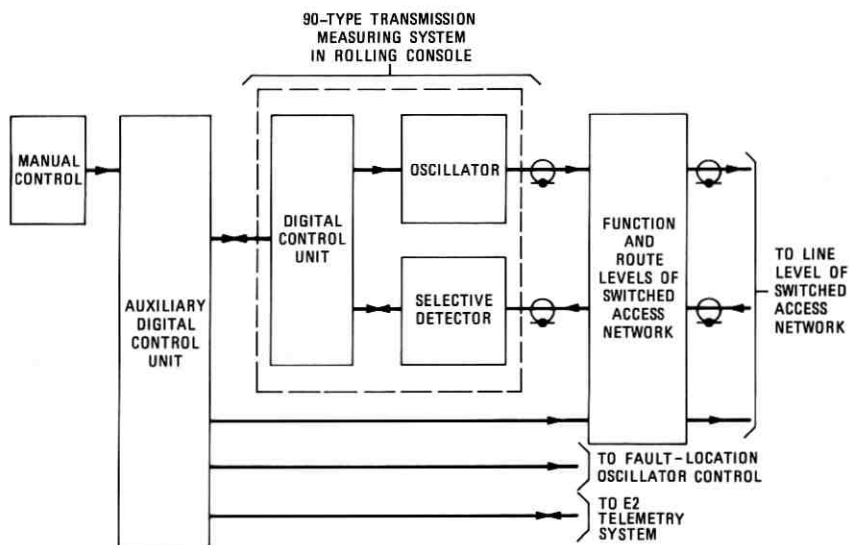

Fig. 5—Transmission surveillance auxiliary.

The TSA operates the local switched access network, fault-location oscillators, and transmission-measuring equipment similarly to the TSC. However, all automatic commands originate at the computer in the TSC and are delivered to the TSA through the E2 data link. When directed by an operational program, the TSA sends raw measured data in digital form through the E2 data link back to the computer in the TSC. The local switched access network and locally powered fault-location oscillators may be operated manually by means of the TSA manual control, which is the same as that in the TSC.

V. SWITCHED ACCESS NETWORK

5.1 Purposes

Signal paths between the measuring equipment and desired L5 system test points are established under computer or local-manual control through a three-level, dual array of multiport, coaxial, ferreed-type switches (Fig. 6). The transmitting array conveys outgoing signals from the oscillator in a TSA or TSC, and the receiving array con-
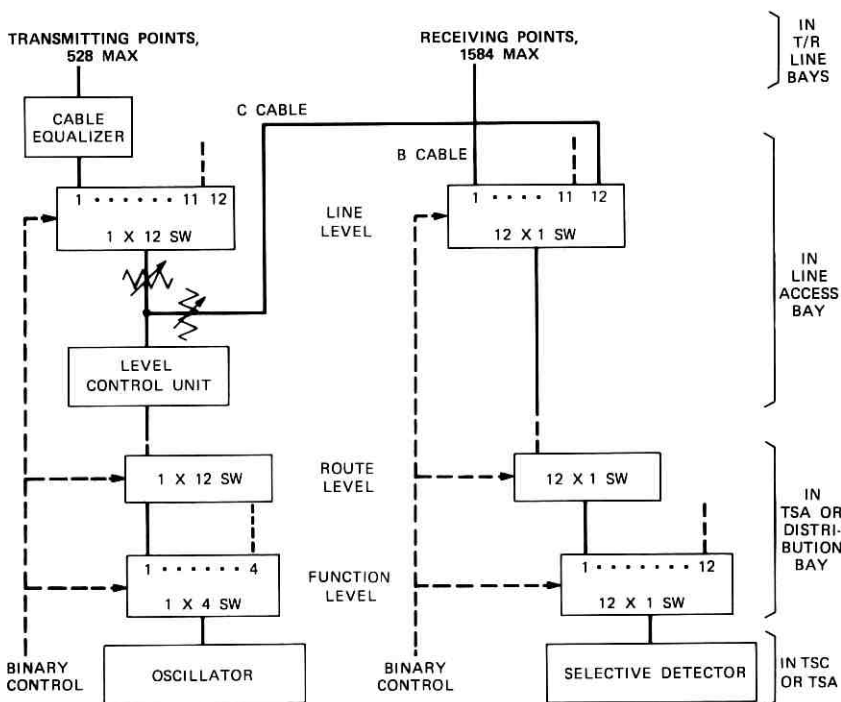

Fig. 6—Switched access network.

veys the signals from certain designated measuring points to the selective detector. Interconnection with the L5 line-protection switching system safeguards working channels from strong test tones.[2] A switched-access-network transmitting path is not completed to an L5 line unless the line is out of service.

5.2 Capacity and growth flexibility

The circuit and equipment groupings of three-level switching (function, route, and line) offer much flexibility and accessing capacity and permit modular growth. The basic arrangement of Fig. 6 yields a maximum of 528 transmitting line-level ports and 1584 receiving line-level ports, not including those used for network calibration. Only 110 transmitting ports and 330 receiving ports are used for switched access to L5 line facilities in a maximally equipped TSC location (10 routes). Most of the remaining capacity will be used for accessing jumbogroup multiplex equipment if the station has much of it.[4]

5.3 Switching-level definitions and equipment placements

A port on the function switch determines the class of an accessed measuring point, such as a particular place in the L5 lines or the JMX equipment. For example, port 1 of the receiving function switch always gets signals from a test point designated receiving line test (RCVG LINE TST), regardless of the selected route or line. A port on the route switch determines the cable entrance, and a port on the line switch determines the coaxial line.

The route- and function-level switches, with an associated power enabler and decoder circuit for controlling all switches, are mounted in the TSA bay at TSA locations and at a TSC in the nearby distribution bay. A line access bay associated with each L5 transmission bay (transmit-receive) lineup contains the line-level switches, which make connections to the L5 line test points. A transmit-receive bay lineup is associated with each cable entrance, which is designated as a route for TSS administration and control purposes. A station without sidelegs and located along a through backbone system has two of these routes defined above, which result from the two cable entrances. Each sideleg contributes one route to a station, since it has one cable entrance.

5.4 Access network control

The switched access network is controlled either automatically or manually by registering specific binary-coded commands in the AUX DCU memory, which outputs this information as continuous dc signals.

The power enabler and decoder circuit decodes and converts these dc signals to decimal arrangements with current capacity and voltage necessary to energize the switch windings. Switch-selection control progresses through the levels as follows:

(i) A signal from the AUX DCU enables the function switch.

(ii) Operating a function-switch crosspoint enables the route switch connected to that crosspoint.

(iii) Operating a route-switch crosspoint enables the line switch connected to that crosspoint.

Interlock logic prevents simultaneous closure of more than one cross-point at the same switching level to assure adequate crosstalk isolation in the access network.

5.5 Optimizing transmission performance of the switched access network

Several circuit features and operating procedures combine to mini-mize the contribution of the switched access network to measuring errors. For receiving measurements, the losses are effectively calibrated out in the paths from the test access points in the transmission bays to the selective detector input. The implementation is as follows:

(i) All B cables are made equal length between the 12×1 line-level switch and the associated receiving measuring points in the transmission bays (Fig. 6).

(ii) The loss-frequency slope from the level-control output through the C cable to the 12×1 line-level switch input is made equal to that of the B cables.

(iii) As part of the measuring program, a reference value is obtained at each frequency by sending a signal from the oscillator, through the level-control unit to the 12×1 line-level switch, and then down through the route and function switches to the selective detector.

(iv) The computer programming takes the reference into account in determining the absolute amplitude of the received signal.

(v) The attenuation in the signal-splitting arrangement at the level-control output brings the calibration-signal amplitudes near those of frequently measured signals to minimize or eliminate attenuator ranging in the selective detector.

For transmitted test signals, the circuit paths from the level-control unit to the transmitting point, where signals are applied to the line, are equalized to a nominal 50-dB flat loss. The accumulation of small errors

in the switched access network, together with the inherent accuracy of the transmission-measuring equipment, should yield an overall measuring accuracy of better than 0.2 dB for received signals. Straightaway measurements should have slightly less accuracy because of switching and measuring equipment involvement at two locations.

VI. FAULT LOCATION

6.1 Procedure

Location of failed or degraded repeaters along an L5 coaxial line is based on measurement of tones transmitted through the line from a four-oscillator fault-location unit associated with each repeater (Fig. 7a). During fault location, dc power and control commands are sent through interstitial wires in the cable from the nearest TSC or TSA to the manhole locations to energize one oscillator group at a time. Two simultaneously closed coaxial switches in an energized oscillator unit connect the four different test frequencies to the input and output of the repeater, as shown in Fig. 7a. The separations between the two low frequencies and between the two high frequencies permit measurement of each individual tone with a selective detector in a distant TSC or TSA.

Before the oscillators are installed in a repeater manhole, the outputs are adjusted so that signals leave the location as follows:

(i) The two low-frequency signals have nominally equal amplitudes if the low-frequency gain of the repeater is normal.

(ii) The two high-frequency signals have nominally equal amplitudes, but not the same as the two low frequencies, if the high-frequency gain of the repeater is normal.

Therefore, the signals received from a basic repeater indicate a gain abnormality if the low-frequency pair or the high-frequency pair differs in amplitude. Signals from a regulating repeater require different interpretation because the gain varies with seasonal temperature changes. Equal amplitudes should be expected in the received signal pairs only when the cable temperature is at the seasonal mean.[5]

6.2 Fault-location-oscillator control and powering

A rectifier power supply and control unit associated with each route, or cable entrance, at a TSC or TSA station directly control one fault-location unit at a time through interstitial wires in the coaxial cable (Fig. 8). In straightaway runs, direct control extends halfway to the adjacent station, which in turn controls the far half of the power
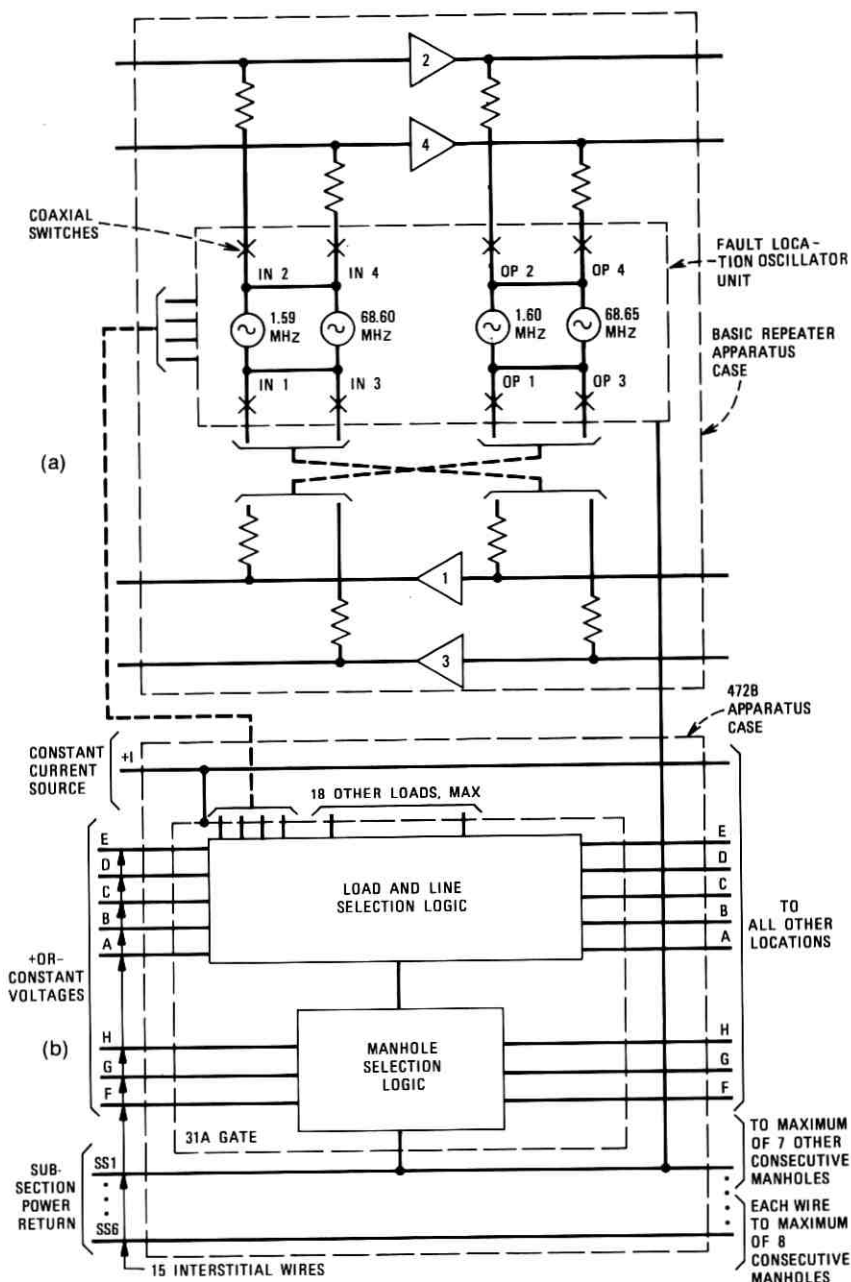
Fig. 7—Fault-location oscillator. (a) Manhole signal arrangement. (b) Manhole control arrangement.
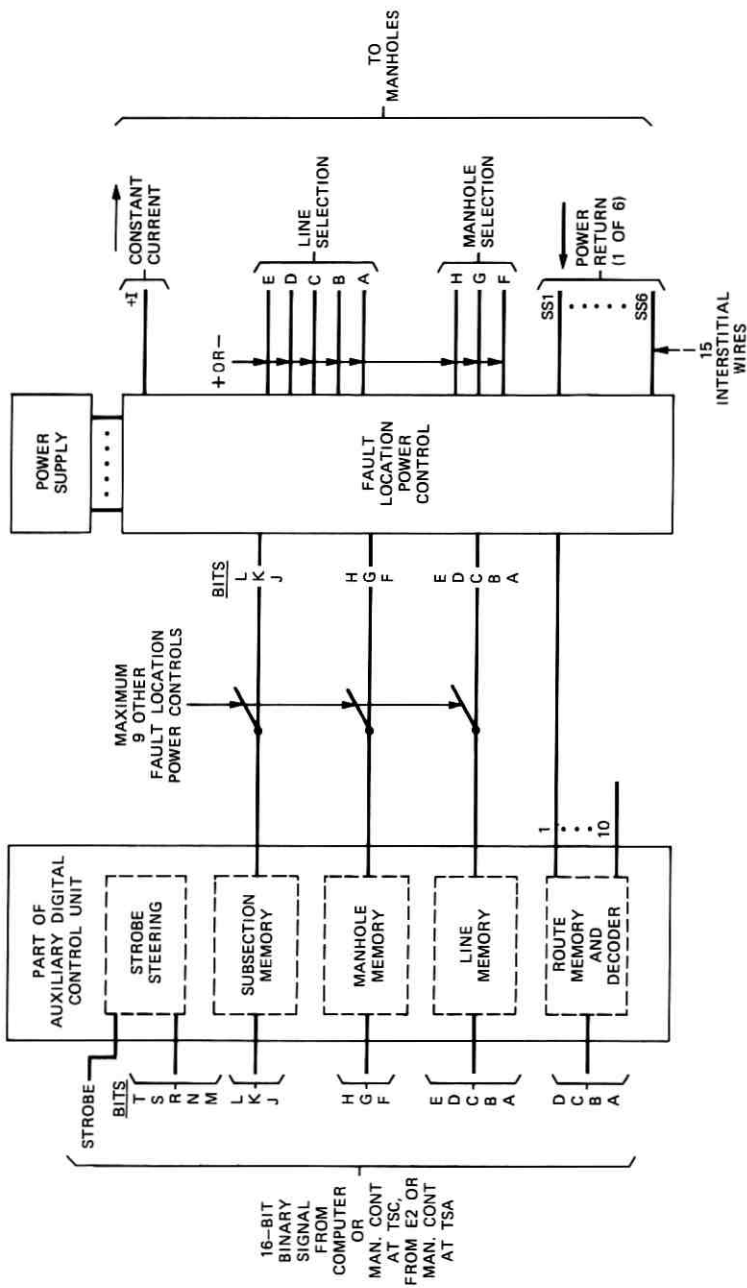
Fig. 8—Station fault-location control.

section. The 48-location capability (maximum) of the fault-location power arrangement permits one-end control in a significant fraction of the sidelegs and end links. Voltage drop in the interstitial wires and the permissible sending-end voltage limit the powering and control distance.

Fault-location-oscillator control typifies the various TSS control operations and is described in some detail as an example. Turning on a fault-location-oscillator unit involves the following sequence at a station (Fig. 8):

(*i*) Upon receiving an appropriate two-word sequence from the computer or manual control at a TSC location, or from the E2 system or manual control at a TSA location, the AUX DCU registers the selection (route, line, subsection, and manhole).

(*ii*) The AUX DCU sends continuous dc voltages through office wiring to operate a fault-location-oscillator power-control circuit.

(*iii*) The operated power-control circuit makes connections to an associated power supply, applies constant current to one interstitial wire $(+I)$, applies binary-coded combinations of plus and minus voltages to eight interstitial wires (A to H), and establishes a connection to one of six subsection wires (SS1 to SS6) that provides a return path for the oscillator-load and control-signal currents.

With one possible exception, each subsection wire serves a group of eight consecutive locations. The last or highest numbered wire serves the most distant group, which may have fewer than eight locations. Most TSS control words use the four most significant bits (N, R, S, and T) of the 16 for strobe steering or addressing. Fault-location-oscillator turn-on also uses a fifth bit (M) for this purpose. The first fault-location control word conveys the route selection via bits A to D; and the second word conveys the selections of line via bits A to E, manhole via F to H, and subsection via J to L. A binary-to-decimal decoder provides a dc signal on one of 10 outputs to enable the power-control circuit associated with the desired route.

6.3 Fault-location power-control circuit

Transistor logic in the enabled power-control circuit operates miniature relays, which apply the appropriate control voltages to outgoing wires A to H and complete connections from the power supply to the $+I$ constant current lead and the desired subsection lead, SS1 to SS6.

The constant-voltage control signals applied to leads A to H are referenced to the subsection leads at the sending end. Positive voltage signifies a logical 1 and negative voltage a logical 0.

6.4 Manhole control logic

At each controlled location, a combinational logic circuit (31A gate) bridged across the interstitial control wires decodes the binary-coded combinations of positive and negative voltages on these leads and performs desired switching (Fig. 7b). The circuit consists of a tree of many simple transistor gates and transistor switches. A mechanical rotary switch assigns one of eight binary codes appearing on wires F, G, and H to each manhole in the subsection. The load- and line-selection logic part of the 31A gate responds to voltages on the A to E leads and selects a desired oscillator load. Coaxial windings in series with the oscillator loads enable directing the outputs to a desired line. Of the 32 binary combinations available with the five A to E bits, 22 only are used for oscillator turn-on and switching, corresponding to the maximum number of coaxials in a cable. Inhibit logic in the power-control circuit at the station prevents sending out an unused code, which could operate the 31A gate ambiguously.

The low current drains in the 31A gate decoding circuits permit individual control of fault-location oscillators at many different locations over a few wires of relatively small diameter. The turned-off locations draw almost negligible current. Since the coded control information is applied continuously while a load is turned on, the interstitial wires need not be loaded or delay equalized as for pulse signals. Use of memoryless logic contributes to reliability by minimizing the possibility of a sustained lockup at a remote point during a malfunction of the control system.

VII. POWER AND FUSING

Commercial 117-volt, 60-cycle power energizes the entire TSS. Should this source fail, an "essential" supply such as a local engine- or turbine-driven alternator carries the load after a few seconds of interruption. Since the TSS is off-line equipment not in the service transmission paths, the five-second or longer start-up time permitted in an "essential" category supply is tolerable.

Solid-state rectifier supplies provide the nominal 5- and 25-volt dc needed to operate the office logic and control circuits in the TSS. Fuses in the distribution paths to the various circuits provide overcurrent protection and convenient power-disconnect capability for maintenance. A blown fuse operates an alarm relay, which activates the

local office alarms and furnishes an indication to the E2 Status Reporting and Control System for transmission to the E2 central.

VIII. FUNCTIONAL ARRANGEMENTS

The TSC equipment is assembled in a desk-type console having a writing area and a turret-type structure with three vertical panel regions just beyond the writing area (Fig. 4). Frequently used units have front-panel access. The left turret has the computer, the center turret the manual control panel, and the right turret the transmission-measuring system. The punched-tape reader is accessible in the left front, behind a hinged door under the writing area. The teletypewriter console is conveniently located to the left of the TSC operator position. Infrequently accessed units such as logic circuits and power supplies are mounted in the rear of the console behind hinged doors. Slides on the logic unit shelves facilitate access for maintenance.

A surveillance distribution bay located near the TSC console contains the function- and route-level coaxial switches for the switched access network, the associated control circuitry, and terminal blocks for connecting numerous station control wires to the TSC and line access bays.

The TSA bay contains essentially all the equipment of both the TSC console and the surveillance distribution bay except the computer, the punched-tape reader, and the transmission-measuring system. Personnel normally stand while operating TSA equipment locally. The same type of transmission-measuring system as that in a TSC is provided in a rolling console.

One line-access bay is associated with each route or cable entrance. It is located in the associated lineup of L5 transmit-receive bays and contains principally fault-location power and control equipment, the line-level switches and level-control unit of the switched access network, and the 31A gate for controlling fault-location oscillators in the station.

Except for the 31A gate, which uses discrete solid-state components, the various logic control circuits throughout the TSS contain mostly commercial integrated circuits assembled on plug-in printed wiring boards. Most mounting frameworks and panels are fabricated sheet aluminum.

IX. SOFTWARE CONSIDERATIONS

Software has been developed to exploit the computational and control facilities of the TSS computer (Section 3.1). Features are provided to minimize operator interaction and to process and present

results in an easy-to-use format. Discussion of the software has been divided into the following categories.

(*i*) Operating system.
(*ii*) Application programs.
(*iii*) Data-base generation and administration.
(*iv*) Diagnostic system.

Computer programming is done in two languages, Assembly and FORTRAN. Assembly language is used primarily to write software drivers (routines to control peripheral devices), where the pseudo-machine-language code is needed to set up and complete input/output operations. The higher-level FORTRAN language is used for applications programs (complete routines for each surveillance activity), where advantage is taken of its speed and flexibility in program development.

A teletypewriter is used to enter system commands and program parameters. The teletypewriter is vendor-modified to allow bit-parallel transfer of information and independent control of the input, print, and punch functions.

The fault-location subsystem, switched access network, and test set are controlled directly by the computer at TSC locations. The computer remotely controls identical equipment at TSA locations via the E2 telemetry facilities.

To relieve the TSC operator of much punched-tape handling, a cassette-type digital magnetic tape unit is also interfaced with the computer; and programs for surveillance system operations are stored on magnetic tape. The magnetic tape unit can read or write on a data track or an address track and access tape files quickly. Tape positioning addresses are prerecorded on the address track of each cassette in proportion to the number of spindle revolutions. To access a file, the tape unit counts spindle revolutions as the file approaches at high speed and then reads the address track only during the last few revolutions to verify that the desired tape position has been reached.

A 64-word program, used to load other programs into the computer via the high-speed punched-tape reader, resides in a protected area of computer memory. This program is used, for example, to load the TSS bootstrap program, which, in turn, loads the operating system from magnetic tape.

The next few sections describe in more detail some programs provided with the TSC. Typical TSS software is given in Table I. Not discussed here are other programs, such as the editor, compiler, relo-

Table I — Typical TSS software

Operating System

Basic Binary Loader
TSS Bootstrap
Directive Processor
Input/Output Control
Drivers for:
  Teletypewriter
  Paper Tape Reader
  Magnetic Tape Unit
  Test Set
  Fault Location Oscillator
  Switched Access Network
  E2 Telemetry
Formatters for:
  Teletypewriter
  Magnetic Tape Unit
Arithmetic Routines

Applications

Pilot Measurement
Fault Location
Line Measurement
JMX Measurement
Surveillance Library
General Purpose Measurements

Data Base

Generators for:
  Pilot Measurement
  Fault Location
  Line Measurement
  JMX Measurement
  TSS Diagnostic

Diagnostic

TSS Diagnostic
Teletypewriter
Paper Tape Reader
Magnetic Tape Unit
E2 Telemetry
Digital Control Unit
AUX DCU
Computer Diagnostic for:
  Computer Instructions
  Computer Memory
  Hardware Functions

catable loader, and magnetic-tape storage programs, used to prepare and manufacture those programs that are a part of the TSC.

9.1 Operating system

The TSS operating system processes a small set of operator directives (Table II); and, to economize magnetic-tape space and reading time, it contains a set of subroutines expected to be used by all applications programs.

Table II — Operator directives

Command	Function
LOAD (NO.)	To recall and start an applications program stored on magnetic tape.
BEGIN	To restart the program currently in computer core.
RUN	To continue a program after a programmed pause for manual operations, or
	To retry certain input/output operations resulting in error messages.
REWIND	To rewind the magnetic tape cassette to clear leader.

Applications programs are stored on magnetic tape and can be recalled by file number into the computer for execution upon an operator-initiated LOAD (file number) command. At the termination of a program or after a programmed pause, the operator may request another program to be loaded and automatically started. He may restart the program currently in the computer core with the BEGIN command; or, in the case of a programmed pause for manual operations, he may cause the current program to continue with the RUN command. Also, using the RUN command, he may request retries on certain input/output operations that result in error messages. All commands for these operations are given by the operator at the teletypewriter.

Vendor-supplied, interrupt-compatible software drivers for the teletypewriter and magnetic tape and an input/output control program for policing requests to these devices are required to perform the operator directives. Software drivers have been developed by Bell Laboratories to control the E2 telemetry system, the fault-location subsystem, the switched access network, and the programmable test sets.

Vendor-supplied library routines such as a formatter for teletypewriter operations and arithmetic subroutines are included in the operating system because of frequent use by applications programs. Also included are Bell Laboratories developed FORTRAN-callable routines for teletypewriter plotting of measurement results and for controlling magnetic-tape operations. The last-mentioned routine permits reference measurement files and data-base files to be stored on magnetic tape and/or recalled under program control.

All the above routines are stored on magnetic tape as file number 2, which is automatically recalled by loading a short paper tape referred to as the TSS bootstrap. The teletypewriter prints the message *WHAT? when the computer is ready to accept an operator directive. At that time, the operator can request, for example, that a specific application program be loaded from magnetic tape and executed. File number 1 is reserved for a tape address directory, which contains the tape position of up to 67 other files for application programs, reference-measurement files, and data-base files.

9.2 Application programs

Use of a programmable computer permits quick and easy implementation of various maintenance activities. In general, an applications program is a complete sequence of computer instructions designed to monitor and evaluate a particular function of the L5 transmission

equipment. For this purpose, the surveillance software includes the list of programs described in the succeeding sections.

9.2.1 Surveillance library

The surveillance library serves as a memory jog for the operator. It prints a list of program names and numbers and optionally will print detailed instructions regarding a specific program.

9.2.2 Automated fault location

After starting the automated fault-location program, the operator enters the set of cable routes (one or more power-feed sections), man-holes, and the lines containing the repeaters to be interrogated. The computer recalls a data base from magnetic tape and checks for the presence of requested repeaters.

For each section of coaxial line requested, the computer initiates the following: (i) calibration of office trunks and coaxial switches in the line-access network between the RCVG LINE TST point (Figs. 6 and 9) and the receiver, (ii) access of the test point, (iii) turn-on of the oscillators for one repeater at a time, and (iv) measurement of the received set of tones. Repeater gain deviations, as derived from tone measurements, are compared with allowed limits and are printed only for out-of-limit repeaters unless the operator specifies, when starting the program, that all measured and derived data be printed. As each section of line is completed, the program releases the switched access network and the fault-location oscillators.

Figure 10 is a sample printout depicting a repeater in trouble. The teletypewriter bell and special characters in the printout provide audible and visual indications of anomalies and out-of-limit conditions. The operator may optionally obtain a teletypewriter plot of the received tone amplitudes as a function of distance along the cable.

9.2.3 Pilot measurement

In a composite message signal, pilot tones may be monitored by the pilot measurement program. The operator may select from the set of line, jumbogroup, mastergroup, and supergroup pilots. Pilots can automatically be measured at both ends of all JMX sections or optionally on a set of lines between two stations. The computer recalls a data base from magnetic tape and checks for allowable combinations of transmitting and receiving stations.

For each section requested, calibration of office trunks and switches is performed at both stations. At the transmitting end of a section, pilots are accessed and measured via the TRMTG LINE TST point (Fig. 9)
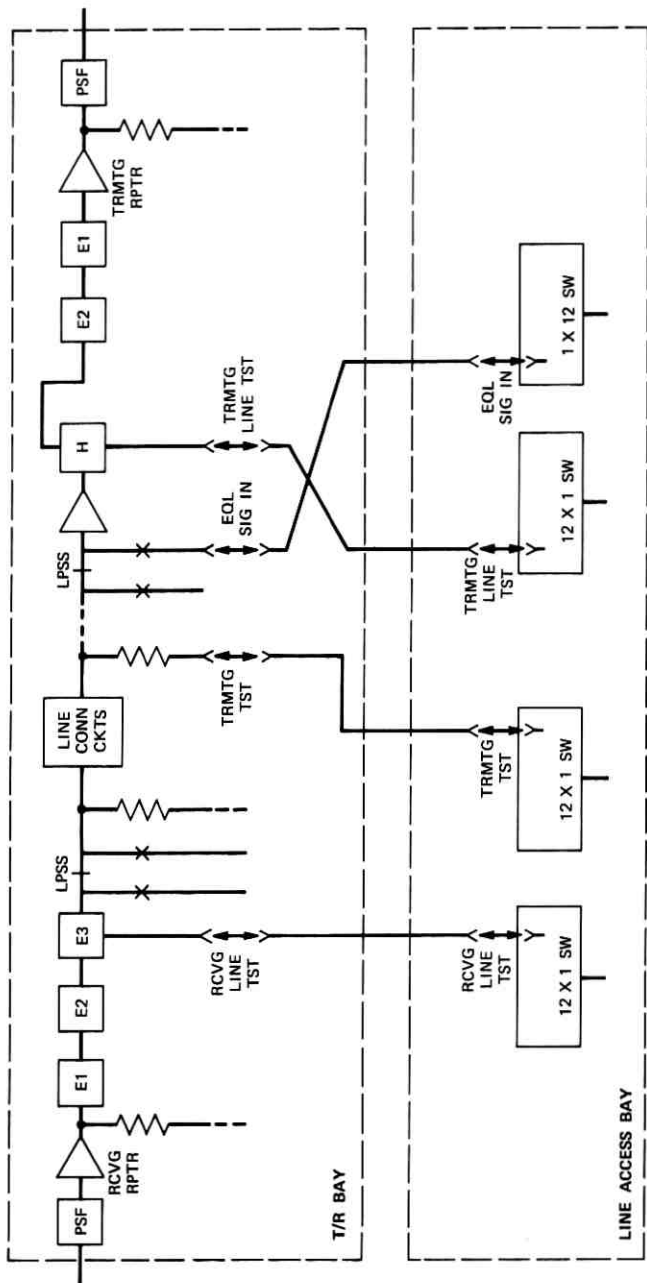
Fig. 9—Switched access network test points.

```
HEADING: DATE, TIME--? 11/28/73 10:15
ROUTE--? 0935
MANHOLES: FIRST, LAST--? 140 250
LINES: FIRST, LAST--? 501
```

```
ROUTE   935     LINE 501
11/28/73 10:15
```

MAN HOLE	1590. INPUT	1600. OUTPUT	LOW DIFF	OUT OF LIMITS	HIGH DIFF	68600. INPUT	68650. OUTPUT
140	-53.12	-53.12	.00		-.62	-53.39	-52.77
150	-53.58	-53.35	-.23		-.27	-53.93	-53.66
160	-53.43	-53.39	-.04		-.05	-54.10	-54.05
170	-53.37	-53.45	.08		-.31	-53.97	-53.66
180	-53.38	-53.72	.34		.38	-53.90	-54.28
190	-53.73	-53.78	.05		-.18	-55.02	-54.84
200	-53.27	-54.04	.77	***	4.23	-49.89	-54.12
210	-53.58	-53.67	.09		.00	-50.19	-50.19
220	-53.74	-53.74	.00		.39	-50.06	-50.45
230	-53.92	-53.96	.04		.50	-50.09	-50.59
240	-53.96	-53.92	-.04		.27	-50.51	-50.78
250	-54.03	-54.24	.21		-.07	-50.68	-50.61

*WHAT?

Fig. 10—Sample printout for automated fault-location program.

and, at the receiving end, via the RCVG LINE TST point. The difference in pilot amplitudes between the two stations is a measure of the gain characteristics of the line. Out-of-limit conditions or, optionally, all pilot amplitudes and differences are printed as a function of frequency. This printout can serve as an early indication of the need to re-equalize a line.

9.2.4 Line measurement

The line measurement program is used to determine the gain-frequency characteristic of a coaxial line with better accuracy and finer granularity than with pilot measurements. A sequence of tones is injected at the EQL SIG IN point in the station at the transmitting end of the desired line and measured at the RCVG LINE TST point in another station. Using the line-protection switching system, the operator must first take the line out of service and operate the message cut-off switch.

After taking the line out of service, the operator selects the transmitting and receiving stations, the line, and the set of frequencies, and

decides whether or not a teletypewriter plot of the gain characteristic is required. Careful study of the printout can be used to evaluate the need for or results of equalizer adjustment. Measurements can be stored on magnetic tape for future comparison with historical data.

9.2.5 JMX measurement

Using the JMX measurement program, the operator can access the remote test switch for each jumbogroup and calibrate office trunks and coaxial switches. The program then measures carrier or jumbogroup and mastergroup pilots at 11 test points per jumbogroup. The data are compared with limits and with historical data. The historical data are updated on magnetic tape, and a printout of out-of-limit conditions is made.

9.2.6 General-purpose loss measurement

The general-purpose loss measurement program operates the transmission-measuring system in a TSC or a distant TSA under computer control to observe directly connected equipment or facilities in the selected station. Loss vs. frequency characteristics are printed and, optionally, plotted.

The operator can select the station address of the test set and all programmable functions of the test set, including the transmit amplitude and set of frequencies. Measurement averaging and reuse of calibration measurements are available.

9.3 Data-base generation and administration

Several application programs use data bases that are prepared by the data-base-generator program and stored on magnetic tape. The data bases determine the arrangement and extent of surveillance equipment to be controlled by a center location, as well as testing limits, etc. Separate data-base files are created for the automated fault location and JMX measurement programs, whereas the pilot and line measurement programs share a data-base file.

Each data-base-generator program is stored on magnetic tape and recalled by the operator using the same procedure as for other applications programs. The program simply asks questions that the operator must answer. For many questions, the operator may elect to use the preprogrammed answers. Questions are answered to define and correlate manhole designations, route numbers, coaxial lines equipped, telemetry addresses, the allowable range of measurement levels, test points available, etc. When all questions are answered, a separate data-base file containing this information is automatically created on

magnetic tape. The operator may rerun the data-base-generator program at any time to verify or review his work by selecting a printout of the contents of the data-base file.

When new or additional equipment must be accounted for in the data-base file, area engineers provide the craft operator with completed forms for updating those files.

9.4 Diagnostic system

A separate set of programs, stored on a second magnetic tape, is dedicated to checking transmission-surveillance-related equipment for proper operation. These diagnostics are called into the computer using the TSS operating system and, if no errors are detected, control is returned to the operating system. Errors are indicated by a computer halt or message on the teletypewriter.

Separate programs are used to test the teletypewriter, high-speed punched-tape reader, digital control unit, and AUX DCU. The magnetic tape unit diagnostic is loaded from punched tape.

The computer diagnostic control program supervises the running of a large set of vendor-supplied computer diagnostics. There are diagnostics for testing all instruction types, core memory, and other computer hardware. Each computer diagnostic has an overlay associated with it to allow nonstop testing and multiple execution of each diagnostic. The control program interacts with the operator to determine the diagnostics to be run.

The TSS diagnostic program provides an indication of the health of the entire surveillance system by exercising all the hardware at the TSC and the TSA's. To narrow down possible trouble causes, each new phase of testing involves a minimum of previously untested equipment. When a problem is detected, more detailed testing is performed in the area affected. To determine what facilities are equipped in each office, the program shares a data base with the automated fault location program.

The degree to which the diagnostic system localizes troubles is established principally by the number and significance of the places where the statuses of transmission and logic signals can be determined automatically and remotely. The TSS automated diagnostic procedures generally localize a fault to a subsystem small enough for knowledgeable craft personnel to find the actual fault quickly.

X. L5 MAINTENANCE PLAN

Since the transmission surveillance system provides the potential for maintaining the L5 system on a network basis rather than as a

series of independent stations, many somewhat complex and inter-related factors must be considered in the evolution of an overall maintenance plan to effectively realize this advantage. To begin with, a TSS controlled by a single TSC should be assigned to a portion of L5 route on a basis compatible with the overall administration of the system. The association of surveillance operations and alarm reporting, as provided by the E2 system, is another important consideration. A third consideration is that of defining the boundary between adjacent surveillance systems along a route. Finally, in developing an overall maintenance plan, intervals for performing routine tasks must be determined by carefully considering such factors as system size, measurement speed, need for update, and output requirements. Of course, considerable blocks of free time must be left for performing demand-type measurements to locate a trouble or to characterize performance, as a result of an external stimulus such as a line switch, E2 alarm, or a line-equalization adjustment.

10.1 Administration aspects

The rules for the layout of a transmission surveillance system are determined as much by administrative and management aspects as by any hardware constraints. Presently, the E2 data facility has a capacity of 16 remote stations, which limits a TSS to 15 TSA stations and one controlling TSC station. The resulting system bound of this number of stations, likely near 800 to 900 route miles, probably approaches the limit that can be monitored effectively by a single TSC. Other considerations include: (i) AT&T-Long Lines area boundaries, (ii) planned E2 alarm reporting and other maintenance arrangements, and (iii) location of strategic stations along the route.

Once a TSS arrangement has been established within a region, it is imperative that a single E2 data facility interconnect all the stations served by the TSS to provide data-transfer capability to and from the TSC. Generally, this data facility will be time-shared with normal E2 alarm polling operations.

Judicious TSC placement enhances the effectiveness of a TSS. Several factors should be considered in selecting the station. It should (i) be on the associated L5 route, (ii) be a major, manned facility, (iii) have significance in the area maintenance plan, and (iv) be a junction point with signal processing (multiplex). Examination of the TSS area and consideration of these points should lead to an appropriate TSC placement. Another highly desirable goal is to colocate the TSC and E2 central, if at all possible, since this would centralize all major maintenance operations. It is important that the TSC location be manned 24

hours a day. Although it is possible to control the TSC computer over a data set/TTY link from another office, perhaps off the route, it is not considered desirable unless craft personnel thoroughly familiar with L5 operations are available to interpret performance data.

10.2 Measurement routines

The criteria for obtaining a viable TSS layout have been given (Section 10.1), and, if availability of computer software is assumed, the role of the surveillance system in the overall maintenance scheme for the L5 system can be defined. As previously discussed, the TSS performs transmission measurements for the purposes of maintaining a high level of performance and locating troubles. These operations fall into two categories:

(*i*) Demand-type measurements.
(*ii*) Routines.

The demand measurements are unscheduled and the need is generally spontaneous, resulting from a trouble condition. The routine operations, where the words *transmission surveillance* apply, are scheduled to realize continued monitoring of performance and to spot any deteriorating conditions. Procedures and associated software for an initial L5 maintenance plan have been based on experience gained during the L5 field trial, in the early stages of the initial system turn-up, and in discussions with AT&T personnel. This plan, when implemented, will assist in maintaining a high quality of service on the L5 system. In simple terms, the plan consists of:

(*i*) Pilot measurements: Transmitted and received line and master-group pilots will be measured daily between all signal-processing points (terminal stations) to characterize the transmission performance of the coaxial lines.
(*ii*) Fault location runs: All line repeaters will be checked weekly, and out-of-limit conditions and regulating repeater gain changes will be printed out.
(*iii*) Jumbogroup multiplex measurements: Test points in all JMX equipment will be monitored on a weekly basis and out-of-limit conditions will be printed out.
(*iv*) Line gain/frequency measurements: Detailed measurements, on an out-of-service basis, of the frequency characteristic of a coaxial line will be made following any repair or realignment operation to verify proper performance.

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(v) TSS diagnostics: Proper operation of the surveillance system itself will be routinely checked on a weekly basis. This includes an overall test at each station to pinpoint any problems in the access network, control circuits, or test equipment as well as diagnostic tests of the computer and associated peripherals at the TSC.

According to calculations for the 815-route-mile initial system between Lillyville, Pa., and Hillsboro, Mo., these routines should use about 50 percent of the total hours in a seven-day week when all 22 tubes are in service. The calculations have assumed that only a few pilot measurements are needed beyond those in the daily overall check between terminal stations to associate a trouble with a particular power-feed section. Methods are being investigated to reduce the measurement times before systems reach full capacity so as to free the TSC for troubleshooting operations when necessary. Furthermore, the presently provided facilities are probably just the beginning of auto-mated transmission-measuring capability. As experience is gained with centralized and automated maintenance and more sophisticated test equipment becomes available, the role of the TSS will grow—perhaps into the areas of acceptance testing, trouble shooting of individual equipment units, and interfacing with other evolving maintenance systems.

XI. ACKNOWLEDGMENTS

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L5 SYSTEM:

Jumbogroup Multiplex Terminal

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The jumbogroup multiplex (JMX) translates basic jumbogroup signals to and from the L5 line spectrum. Each of three jumbogroups provides 3600 4-kHz voice circuits. In addition to frequency translation, the JMX provides regulation and equalization on a jumbogroup basis and automatic one-for-one protection.

Performance objectives are discussed and interpreted from the perspective of their influence on the design philosophy and approach used and the resultant physical realization. All subsystems are presented in detail. Particular mention is made of the double-balanced diode-ring modulator developed for JMX application.

I. INTRODUCTION

While gross performance objectives for analog multiplex equipment are available, the detailed statement of design objectives is part of the development task. The basic guiding principle is that the resultant system, the aggregate of line and terminal equipment, must provide transmission paths of quality and reliability appropriate to the Bell System communication network, and it must do so at a cost sufficiently lower than other alternatives to warrant development by Bell Laboratories, manufacture by Western Electric, and purchase by Long Lines and the operating telephone companies.

The basic function of the jumbogroup multiplex (JMX) terminal is to translate basic jumbogroup signals to and from the L5 line spectrum. Each of the three jumbogroups is composed of six mastergroups. These signals must not be corrupted or distorted by the JMX in a manner or degree that would compromise the quality of transmission. The equipment must be reliable and maintainable. Derivable from these considerations are a host of performance objectives, including those discussed in the following sections.

1.1 Terminal noise

The overall objective on noise, including thermal noise and intermodulation, for a 4000-mile, transcontinental, L5 system is 40 dBrnc0.[1] Of the total, 39.4 dBrnc0 is allocated to the L5 line, and the remaining 31.2 dBrnc0 is allocated to the aggregate of all terminal equipment. While the L5 line is, in a sense, a totally new design entity, such is not the case in the associated terminal equipment. The JMX has to share the terminal noise allocation with an existing hierarchy of equipment, including channel, group, supergroup, and mastergroup banks. Consider a representative transcontinental connection including two channel banks, five supergroup multiplex (LMX) terminals, seven mastergroup multiplex (MMX) terminals, and eight JMX terminals. Assuming 18 dBrnc0 performance for the LMX and MMX terminals and 10 dBrnc0 for the channel banks, a maximal allocation of 18 dBrnc0 results for the JMX. Allowing 1 dB for misalignment and other vagaries, including the possibility of other connections more demanding than the one cited, the JMX design objective was set at 17 dBrnc0, with the realization that an even lower number would be desirable. In essence, it appeared that at least initially the design philosophy should be to obtain the lowest noise possible.

To put the 17-dBrnc0 objective into perspective, it is helpful to view the JMX challenge relative to the most comparable multiplex available, the MMX-2.[2] In an optimally designed multiplex terminal, the controlling noise sources are the modulators and the associated amplifiers following the modulators. Normally, the intermodulation distortion of the modulators is greater than that of the other transmission apparatus, so that lowest signal levels are found at modulator outputs. Since levels are low at such points, the noise figure of the following amplifier is of greatest importance. Levels are chosen to minimize the combined noise from these critical nodes. Since the JMX uses multiple steps of modulation and demodulation, while MMX-2 uses single steps, the JMX has twice as many critical nodes in each transmission path as the MMX. Furthermore, the bandwidth of the jumbogroup signal is six times that of the mastergroup signal. For third-order intermodulation distortion alone, the number of products the JMX must contend with is 16 dB greater than in the MMX. In a sense, JMX must be many decibels better than MMX-2 to yield comparable noise performance.

1.2 Crosstalk

Based on studies of typical system configurations, with the multiplicity of occasions for crosstalk, an objective of 85 dB equal-level

coupling loss (ELCL) has evolved for analog multiplex equipment. This objective was assumed for the JMX.

The sources of crosstalk in multiplex equipment are many and varied, and include many modes not encountered elsewhere. Crosstalk may occur between different jumbogroup signals, between different portions of the same jumbogroup, between signals at different stages in the modulation-demodulation process, and in an intelligible or noise-like fashion. Coupling may exist between signals, between carriers, between pilots and carriers, etc. While an exhaustive discussion of crosstalk is inappropriate here, one typical illustrative example will be mentioned.

Given a carrier with a tone 4 kHz removed from the carrier at a level 79 dB lower than the carrier level, intelligible adjacent channel crosstalk appears in almost every channel at a level approximately 85 dB below the level of the interfered channel. Because of this mode of crosstalk, spectral purity of JMX carriers is important. Since JMX carriers are as high in frequency as 91.648 MHz, filtering, shielding, and grounding are critical. Common ground impedances of very small magnitude are sufficient to cause unacceptable crosstalk performance.

1.3 Spurious tones

Tones falling in a voice channel are particularly annoying. Often, tones generated from several sources add coherently (voltage addition). Because of the multiplicity of sources of tones, the design objective for JMX was that no tone falling in the jumbogroup passband should have a level exceeding -70 dBm0. This becomes particularly challenging when, in the JMX, the source of the tone may be a 15-dBm, 91.648-MHz carrier, and transmission levels are as low as -43 dB. The resulting implications on isolation of separate paths, carrier balance, and filtering of carrier leak signals are significant.

1.4 Frequency offset

The transcontinental objective on frequency offset is less than 2 Hz. This implies an accuracy requirement of one part in 10^8 on the effective carrier frequencies used in the JMX.

1.5 Passband distortion

The overall misalignment objective for a transcontinental L5 connection is less than ± 4 dB. This requires that the JMX passband be flat to ± 0.2 dB. In addition, since the JMX may carry digital signals, delay distortion must be limited.

1.6 Reliability

The L5 line,[3] with its automatic line-protection switching system,[4] is highly reliable. If terminal equipment is not to add noticeably to the mean outage time for a transcontinental connection, based on the system configuration considered earlier, the objective for each JMX path (transmit or receive) becomes less than 0.14 minute per year. This corresponds to an equivalent failure rate of about 200 FIT's (mean time to failure is 500 years), assuming that the mean time for repair is one hour.

1.7 Maintainability

While it is difficult to become quantitative relative to maintainability, several observations of a qualitative nature may be made. Each piece of transmission equipment can handle as many channels as a whole L4 system. The equipment may be housed in unmanned main stations.[5] Many modes of failure are of a subtle nature. It therefore seems desirable to use automatic, remote diagnostics extensively. Routine measurements at the bay should be held to a minimum. Most troubleshooting on site should require no removal of modules prior to trouble isolation. Status indicators should be provided for the controlled and trouble states of the equipment. In short, the equipment should be easy to maintain and difficult to operate incorrectly.

1.8 Cost

The cost of the JMX on a per-channel basis may be reasonably viewed from two perspectives: (i) its contribution to the total terminal cost for the terminal equipment required to take a voice signal from voice frequency to basic jumbogroup frequency and back, including signaling, and (ii) its contribution to the cost of an L5 system for a system length of a few hundred miles or more. It has been estimated that the per-channel cost of the JMX amounts to less than 2 percent of either the terminal or line cost; therefore, the JMX cost would have very little influence on the total cost of providing a voice channel.

II. GENERAL DESCRIPTION

2.1 Transmitting circuits

An overall block diagram of the transmitting arrangement, typical of any jumbogroup, is shown in Fig. 1. Basic jumbogroup signals, composed of six mastergroup signals, and a 5.888-MHz jumbogroup pilot are fed to the JMX from the basic jumbogroup trunk bay (BJGT).[6]
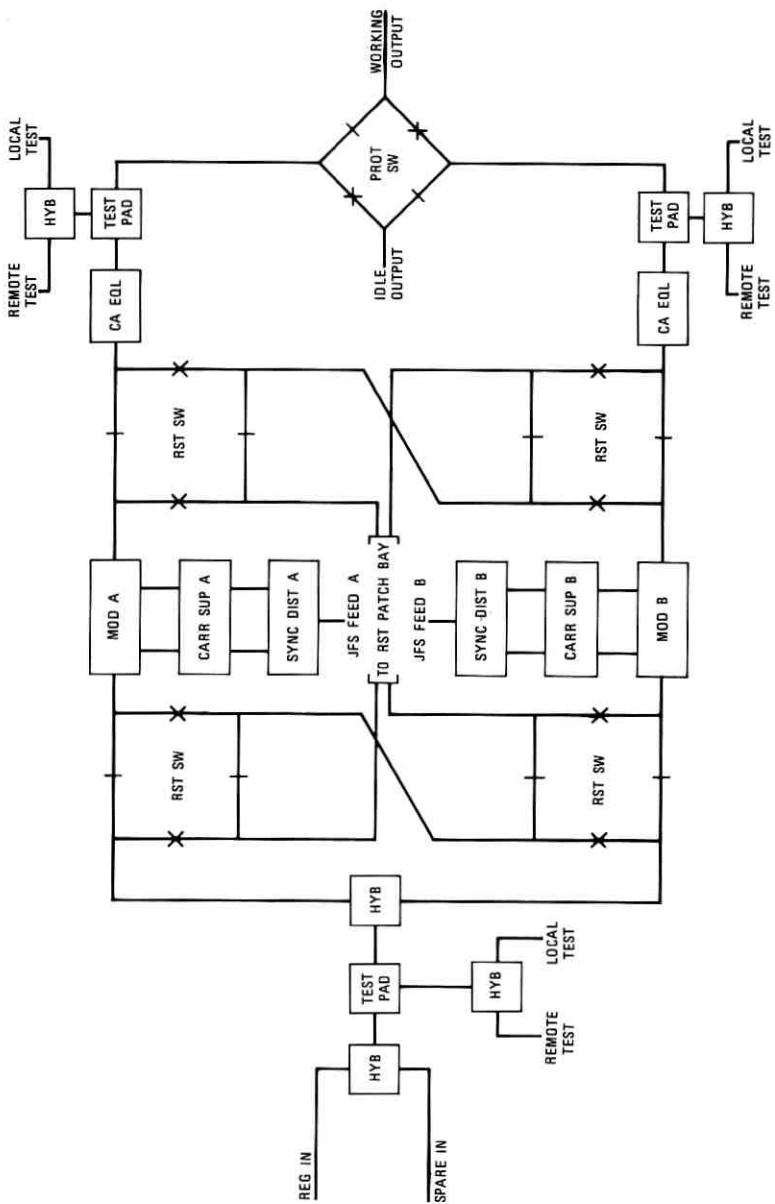
Fig. 1—Jumbogroup transmitting arrangement.

These signals, contained between 0.564 and 17.548 MHz, first enter the JMX at the transmitting jack field. Two input ports are provided. The first or regular input is that through which the signal normally passes. The second or spare input is provided for emergency patching purposes and is normally terminated. These two inputs are combined in an input hybrid transformer and then connected to a test pad that provides a low-loss (0.6 dB) through path and a high-loss (30 dB) bridging path. The high-loss or low-level output is fed to a splitting transformer, and provides a local and a remote test point, both of which are isolated from the transmission path by about 33 dB of loss. Local test points are accessed via jacks mounted at the front of the bay. Remote test points are provided to allow access for automated, centralized test equipment. The low-loss pad output is fed to a splitting hybrid that provides signals to the redundant A and B paths of the transmitting side.

The basic jumbogroup signal then passes through the restoration access switches which, when operated, are capable of providing an appearance of the basic jumbogroup signal and access to the input of the associated modulator at the restoration patch bay. These switches are connected in a manner that allows monitoring of the interbay cabling when the switches are in the unoperated condition.

The basic jumbogroup signal then goes to the modulators, where it is filtered and translated to any one of three jumbogroup line assignments. Multiple steps of modulation are used, and the required carriers are provided by the associated carrier supply. Carrier supplies are redundantly provided for each jumbogroup. Each carrier supply provides the carrier signals for an associated modulator and demodulator. Carriers are generated by appropriately mixing internally generated signals with signals provided by the sync-distribution circuit. There are two sync-distribution circuits per bay, each of which uses redundantly provided jumbogroup frequency supply (JFS)[7] signals to generate a number of highly stable reference signals for transmission to all of the A or B carrier supplies.

The jumbogroup signal, at line frequency, is then fed through the output restoration switch to the cable equalizer. The output restoration switch, connected in a manner similar to that of the input restoration switch, provides access to the modulator output. The cable equalizer, located in the transmitting line-interface unit, compensates for the loss of the cable connecting the JMX to the L5 line bay.[6] Levels are adjusted in the transmitting line interface to provide a stepped preemphasis for transmission over the repeatered L5 line.

The cable equalizer output passes through a test pad which provides a high-loss access for local and remote testing and a low-loss path to the transmitting protection switch. The transmitting switch accepts the A and B output signals as inputs, and provides a working and idle output. The state of the switch, set by the switch-control circuitry, determines which input will be connected to the working and idle outputs. Both outputs are fed to unequal-ratio hybrid transformers which provide a high-loss (7 dB) output for monitoring purposes, and a low-loss (1 dB) output that is connected to the transmitting jack field. The working output signal is then cabled to the L5 line bay, and the idle output, available for emergency patching through spare cabling, is normally terminated.

2.2 Receiving circuit

On the receiving side (Fig. 2), the complete L5 spectrum is redundantly cabled to the receiving jack field of each jumbogroup equipment. The signal first passes through a test pad that provides high-loss access for testing and a low-loss connection to the associated cable equalizer. The cable equalizer compensates for the loss of the cable between the L5 line bay and the JMX. The signal is then fed to an input restoration switch which is capable of providing an appearance of the L5 line signal and access to the demodulator input at the restoration patch bay.

The received signal then goes to the demodulator where the desired jumbogroup is selected and translated to basic jumbogroup frequency. After passing through an output restoration switch, which can be used to gain access to the demodulator output for the restoration patch bay, the signal is fed through a test pad to the basic jumbogroup equalizer, a manually adjustable, multistage equalizer that will provide* the capability of correcting for misalignment accrued over many miles of the repeatered L5 line on a jumbogroup basis. The signal then goes to the regulator which monitors the level of the received jumbogroup pilot and modifies its gain accordingly.

The regulator output is fed through another test pad to the receiving protection switch. Working and idle outputs are connected to unequal-ratio hybrid transformers that provide access points for automatic monitoring in addition to output signals connected to the receiving

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* Space and power have been provided for the jumbogroup equalizer in the JMX bay. Determination of the appropriate characteristics of this equalizer awaits evaluation of misalignments for working L5 systems.
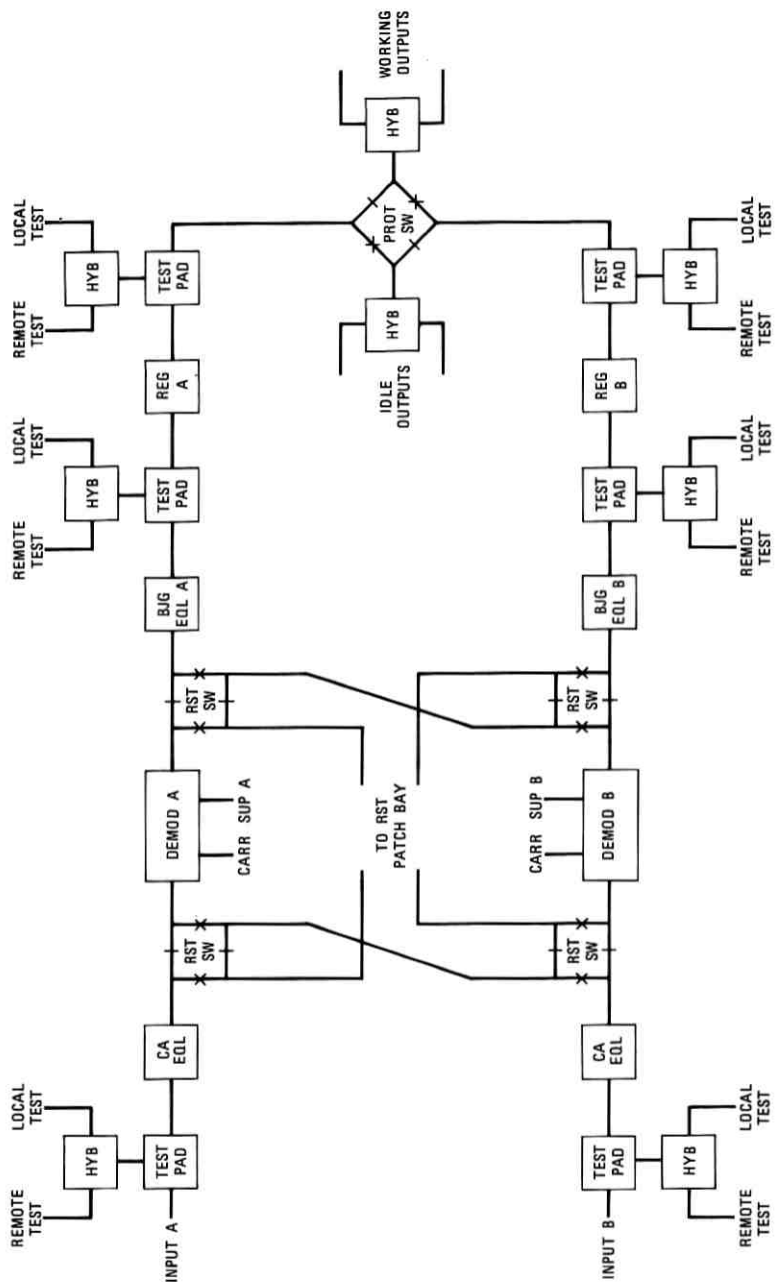
Fig. 2—Jumbogroup receiving arrangement.

patch field. Additional equal-ratio hybrid transformers are used to provide dual appearances of the working and idle output signals. One of the working output signals is cabled to the BJGT. The remaining ports, available for emergency patching, are normally terminated.

III. JMX SUBSYSTEMS

3.1 Modulators and demodulators

The modulator and demodulator subsystems for all three jumbo-groups are shown in Fig. 3; transmission levels are as indicated. In each case, the subsystems used in the A and B sides are identical.

Three types of amplifiers are used in the modulator and demodulator subsystems. The 9-dB fixed-gain amplifiers and the 12 ± 2-dB adjustable amplifiers are transmission quality amplifiers with a controlled transmission band running from 0.5 to 100 MHz. They are realized in the hybrid integrated circuit (HIC) technology. The 9-dB amplifier is designed for low-level application and has a 5-dB noise figure. The 12-dB amplifier has been designed to handle higher signal levels without introducing appreciable intermodulation distortion. Both amplifiers are multistage, major-loop feedback designs with 75-ohm input and output impedances. The 15-dB carrier drive amplifier is designed to provide a 15-dBm carrier signal for the modulator. In addition to providing highly linear gain for carriers up to 91.648 MHz, this amplifier provides at least 40 dB of reverse isolation to control crosstalk that might otherwise be established through the carrier drive path. A multistage local-feedback design with printed-circuit realization is used. All amplifiers use feed-through filters for battery connections.

The 17A modulators are double-balanced diode-ring modulators using Schottky barrier diodes and 75:300-ohm center-tapped transformers.[8] These are highly linear modulators providing at least 40-dB carrier and signal balance. Nominal port impedances are 75 ohms. A single design is used throughout the JMX. More will be said about the 17A modulator in Section 3.16.

The filters[8] have been designed on a system basis; i.e., the loss requirements have been specified in an interactive manner in an attempt to reach a global optimum without placing undue stress on any one design. Often the attenuation required for some undesired sideband can be obtained more easily through the combined effect of two or more filters than in a single filter. The systems of filters were specified to provide combined attenuation to undesired modulation
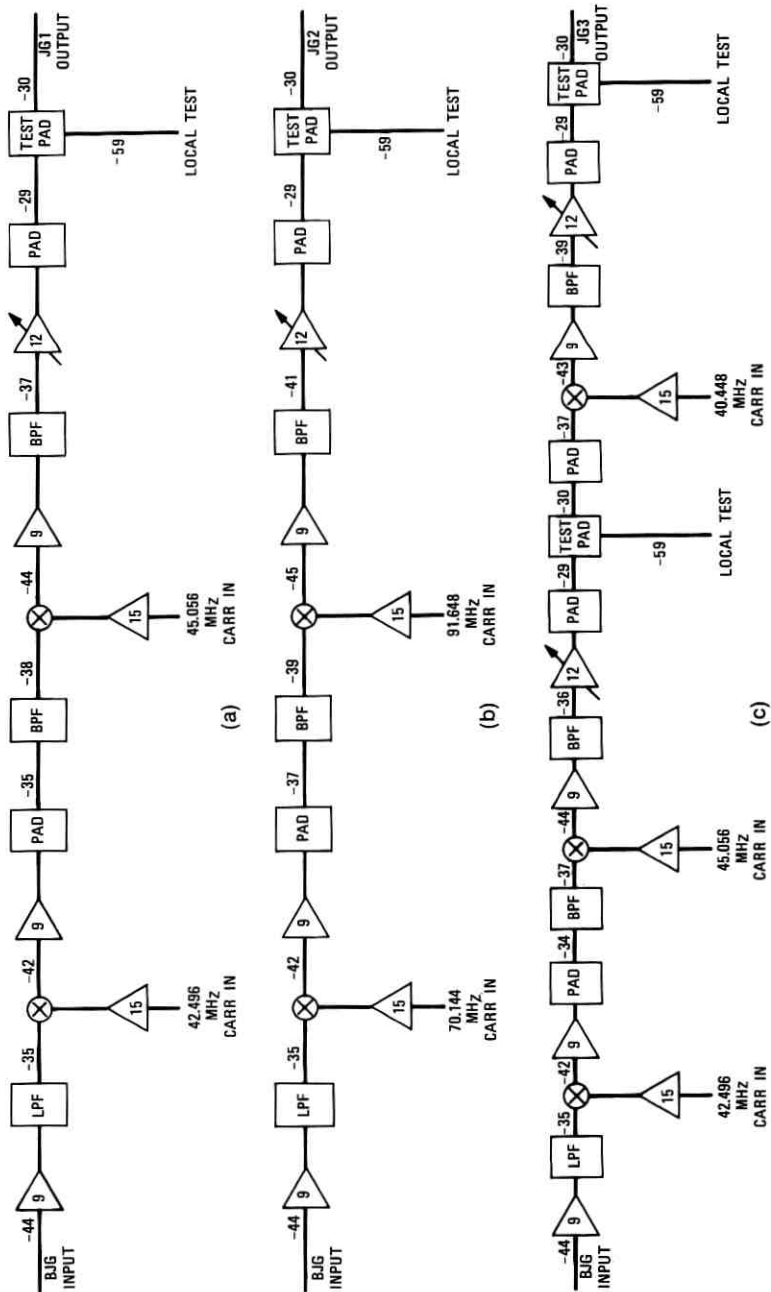
Fig. 3—Modulator and demodulator subsystems for jumbogroups 1, 2, and 3. (a) Jumbogroup 1 modulator. (b) Jumbogroup 2 modulator. (c) Jumbogroup 3 modulator.
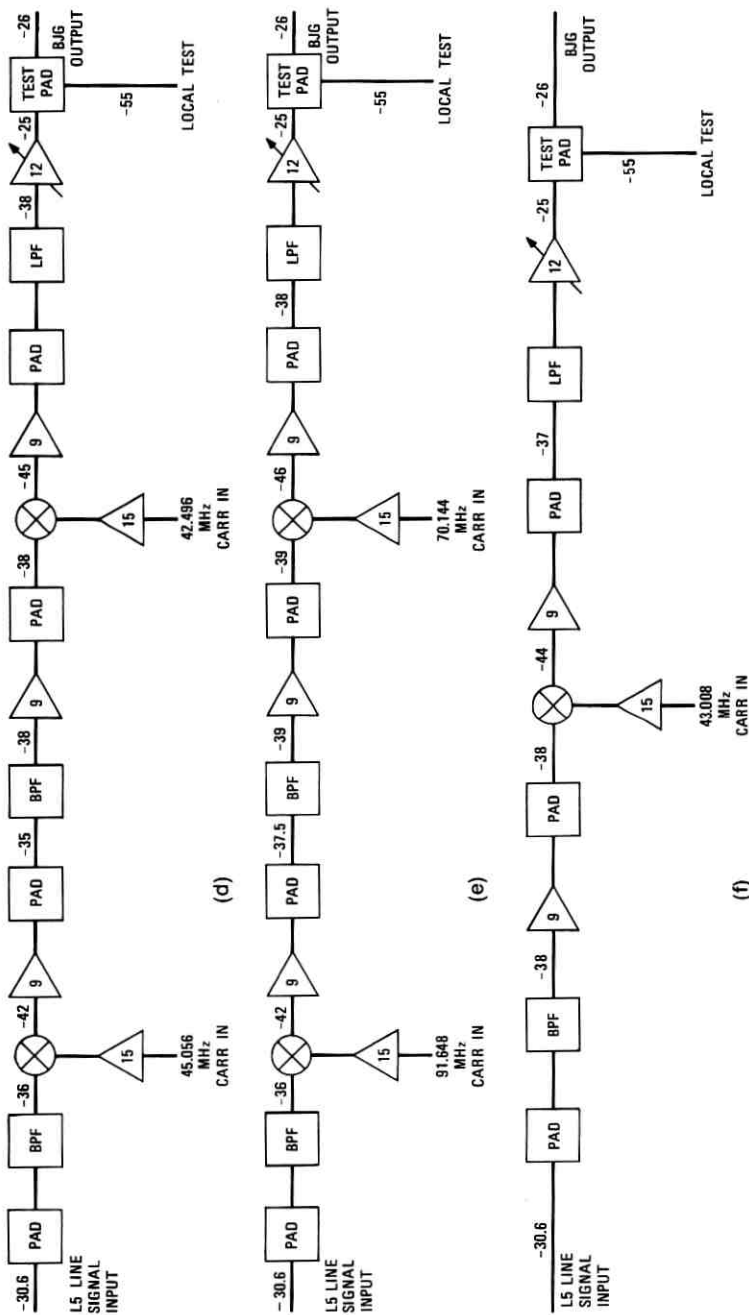
Fig. 3 (cont.)—(d) Jumbogroup 1 demodulator. (e) Jumbogroup 2 demodulator. (f) Jumbogroup 3 demodulator.

products and carrier leak sufficient to keep them at least 85 dB lower than the level of the desired sideband. All filters have nominal input and output passband impedances of 75 ohms.

3.2 Jumbogroup 1 modulator

Ideally, the basic jumbogroup signal fed to the JMX should contain no energy above 17.548 MHz. If some disturbance is present above the desired input spectrum, it could corrupt the message spectrum either by being superimposed as a result of linear transmission through the modulator (signal leak) or by being translated into the message band via the modulation process. Accordingly, after the input signal is raised to the appropriate transmission level via the input amplifier, it is passed through a low-pass filter to reject any spurious high-frequency disturbance. The basic jumbogroup spectrum is then translated in frequency using a 42.496-MHz carrier. The signal is amplified and its level is adjusted to a more optimal value for a subsequent step of modulation. Next, a bandpass filter selects the lower sideband between 29.948 and 41.932 MHz. The signal is modulated once more, using a 45.056-MHz carrier. After amplification, the lower sideband located at 3.124 to 20.108 MHz is selected by a second bandpass filter. The modulator subsystem output level is then achieved through the use of an adjustable 12-dB amplifier, a fixed-loss pad, and a test pad that introduces a 0.6-dB loss in the signal path and provides a bridged test point with 30-dB isolation.

The net frequency shift provided by the jumbogroup 1 modulator is 2.56 MHz. Since the input spectrum overlaps the output spectrum for jumbogroup 1, multiple steps of modulation are mandatory. Signal leak prohibits using one step of modulation (or demodulation) for jumbogroup 1.

3.3 Jumbogroup 2 modulator

Jumbogroup 2 is processed in a manner similar to that for jumbogroup 1. The first step of modulation uses a 70.144-MHz carrier. The first bandpass filter selects the lower sideband between 52.596 and 69.580 MHz. The second carrier frequency is 91.648 MHz, and the jumbogroup 2 signal occupies the band of 22.068 to 39.052 MHz. The net shift for jumbogroup 2 is 21.504 MHz.

It is often more desirable to use the lower sideband, since many of the undesired byproducts of modulation corrupt the upper sideband. In jumbogroup 2, using a single step of modulation and selecting the lower sideband would cause the most difficult filtering—attenuation

of carrier leak and rejection of the upper sideband—to be done in the neighborhood of 40 MHz. Using two stages of modulation greatly facilitates this task by shifting the chore to the region of 20 MHz.

3.4 Jumbogroup 3 modulator

Jumbogroup 3 uses three steps of modulation. The first two steps are identical to those of jumbogroup 1. The third step of modulation uses a 40.448-MHz carrier. The third bandpass filter selects the upper sideband between 43.572 and 60.556 MHz. The net frequency shift is 43.008 MHz.

With either one or two steps of modulation, the attenuation of carrier leak and the rejection of the nearby sideband would require filters of questionable realizability for jumbogroup 3. Thus, three steps of modulation were necessary. The first two steps may be viewed as providing a baseband signal with significant separation between the lowest signal frequency and dc. The result is that adjacent sidebands are separated by about 6 MHz after the third step of modulation. This, coupled with the use of the upper sideband (in this case, no spurious modulation products overlapped the upper sideband), greatly facilitates filter synthesis.

3.5 Jumbogroup 1 demodulator

Following a slight level adjustment to place the signal at a more optimal level for the first step of demodulation, the jumbogroup 1 signal is selected from the L5 line signal by the first bandpass filter whose passband extends from 3.124 to 20.108 MHz. Using a 45.056-MHz carrier, the first step of demodulation shifts the input spectrum up in frequency. Following amplification and level adjustment, the second bandpass filter selects the lower sideband signal between 24.948 and 41.932 MHz. The filter output goes through a 9-dB amplifier-pad combination to provide isolation between the filter and the following modulator. This signal passes through a second stage of demodulation using a 42.496-MHz carrier. The signal is amplified, and the lower sideband signal at basic jumbogroup frequency is selected by the output low-pass filter. The basic jumbogroup signal is further amplified, then fed to a test pad that provides a local bridged test point.

The jumbogroup 1 demodulator is similar in most aspects to the jumbogroup 1 modulator. It uses the same carriers and uses filters with identical passbands (although the detailed nature of the rejection bands is not always the same).

Modulators of the type used in the JMX are bilateral. That is, a signal applied to the output port will be translated in frequency and appear at the input port in the same manner that signals applied to the input are translated and appear at the output. Consider a modulator with conversion loss C terminated on its output by a load whose return loss, with respect to the average modulator output impedance, is R_o, and terminated on its input by a source whose output return loss, relative to the average modulator input impedance, is R_i. Signals appearing at the modulator output will have some energy reflected back through the modulator, and some of this energy will, in turn, be reflected from the source output through the modulator once more. The result will be an echo-like signal superimposed on the original signal at the modulator output. The reflected signal will be attenuated relative to the original signal by $2C + R_o + R_i$ dB.

The modulator input and output impedances are complex and periodically time-varying, and are not capable of being carefully controlled and manipulated without significant decrease in conversion efficiency. In practice, the average input impedance is controlled to some nominal value. For the JMX, this nominal value is 75 ohms. Similarly, amplifiers, pads, and in-band filter impedances are designed to be 75 ohms. Filters are designed to pass in-band energy and reject, principally through reflection, out-of-band energy. Consequently, filters typically have rejection-band terminal impedances corresponding to open or short circuits.

Assume that a modulator is driven from a filter and operates into an amplifier. If the amplifier input impedance has an average return loss of $R_o = 18$ dB, relative to the modulator output impedance, the filter out-of-band return loss is $R_i = 0$ dB, and the conversion loss is $C = 6$ dB, then the reflected signal will be suppressed 30 dB relative to the signal with which it adds, assuming that the signal that appeared at the filter input was an out-of-band signal. The interfering signal will be at the same frequency as the interfered signal and will add on a voltage basis, depending on the phase relationship between the two signals. If the angle is constant across the band, a slight level shift will occur. If the angle is not constant, but changes slowly with frequency, passband distortion, gradually changing with frequency, will be introduced. This can be compensated for through the use of simple deviation equalizers designed to attend to the aggregate distortion introduced by filters and other apparatus. If, however, the angle of the reflected signal changes rapidly, as it would if the angle of the filter output impedance changed in the region of resonance of a

crystal, then sharp, unequalizable notches could be introduced into the jumbogroup passband. The reflected energy could cause variations in passband loss as large as ±0.27 dB for the relative levels indicated above. This was, in fact, experienced in initial measurements of jumbogroups 1, 2, and 3 receiving equipments, and required the introduction of 9-dB pad-amplifier combinations to improve the return loss of driving impedances for selected modulators, as indicated in Fig. 3.

3.6 Jumbogroup 2 demodulator

The jumbogroup 2 demodulator is functionally identical to that of jumbogroup 1. It uses the same carriers and has filters with the same passbands as the jumbogroup 2 modulator.

3.7 Jumbogroup 3 demodulator

The jumbogroup 3 signal is demodulated in a single step; thus, all three jumbogroups traverse four modulators during the modulation/demodulation process. After level adjustment, the jumbogroup 3 signal is selected from the L5 line spectrum by a bandpass filter with a passband between 43.572 and 60.556 MHz. The signal then goes through a 9-dB amplifier and pad to provide isolation between the filter and the following modulator. The jumbogroup 3 signal is then demodulated with a 43.008-MHz carrier and, following amplification and level adjustment, is sent through a low-pass filter to isolate the basic jumbogroup signal. Following additional amplification, the signal passes through a test pad that provides local test access.

3.8 Observations

A major and critical aspect of the design of an FDM terminal involves the ordering of apparatus and the settings of internal levels in the modulator and demodulator subsystems. The output of a double-balanced diode-ring modulator contains the dominant upper and lower sidebands, carrier leak, and a multitude of other signals.[9] The energy in either sideband is of a magnitude comparable to that of the energy in the carrier leak. If the modulator output is fed through a bandpass filter into an amplifier, then the undesired sideband and the carrier leak can be attenuated so that the total signal power carried by the following amplifier is greatly reduced. This tends to reduce the intermodulation distortion generated by the amplifier. It also tends to degrade the effective noise figure of the amplifier by an amount equivalent to the filter passband loss. Furthermore, should the signal

next be fed to a subsequent stage of modulation, the next modulator is presented with a broad band of noise covering not only the message band, but the image band as well. If the order of the amplifier and filter is reversed, then the thermal noise problem is mitigated at the expense of increased intermodulation distortion in the amplifier. The more beneficial arrangement depends upon a number of parameters, including filter loss, amplifier noise and intermodulation distortion, modulator intermodulation distortion, and carrier leak. While preliminary analysis of the gross characteristics of the various transmission elements is helpful in establishing a starting point, the optimal subsystem configuration can only be obtained through extensive, if not exhaustive, examination of the various possible configurations. This has been done for all modulator-demodulator subsystems using noise-loading techniques.[10]

It has been found that the setting of internal levels is equally important. Signal and carrier levels also were optimized using noise loading.

3.9 Carrier generation

Jumbogroup signals are translated to their L5 line-frequency allocation through multiple steps of modulation. The net frequency translation they experience may be viewed as having been achieved through a single modulation step using an equivalent carrier frequency as follows. Consider the first step of modulation to have used a carrier of frequency f_1. Following the first step of modulation, a signal of frequency f_s would be translated to $f_1 - f_s$. The second step of modulation, using carrier frequency f_2, would translate the signal to $f_2 - f_1 + f_s$. The net translation is $f_e = f_2 - f_1$. In jumbogroups 1 and 2, where two steps of modulation are used, the above applies. For jumbogroup 3, a third modulation step, using a third carrier frequency f_3, is employed. In this case, the equivalent carrier is $f_e = f_3 + f_2 - f_1$. To demodulate the jumbogroup signals accurately, only the equivalent carrier frequencies need be generated at the receiver. This principle is used extensively in the JMX. The equivalent carrier frequencies are 2.560, 21.504, and 43.008 MHz for jumbogroups 1, 2, and 3, respectively. The frequency stability of these frequencies is determined solely by the stability of the jumbogroup frequency supply (JFS).[7]

The JFS generates three reference signals at frequencies of 20.480, 2.560, and 1.024 MHz. The second and third reference signals are obtained from the first through division of a 20.480-MHz signal by 8

and 20, respectively. These signals are further processed by the sync-distribution circuits to yield the 2.560- and 21.504-MHz signals which are fed to the carrier supplies. The carrier supplies, using the appropriate input signals and locally generated signals, generate the carriers for modulation and demodulation. The functional separation between the JFS and sync-distribution circuit is somewhat arbitrary. The counters that provide the division by 8 and 20 operations could have been located in the sync-distribution circuit, or much if not all of the sync-distribution circuit could have been housed in the JFS. Since it may be feeding up to 20 JMX bays, failure of the JFS could be catastrophic; therefore, the JFS should be as uncomplicated as possible, with every effort made to maximize its reliability. The only exception to this rule was the inclusion of the count-down circuitry in the JFS. These counters are implemented using high-speed emitter-coupled logic. It was expected that locating such circuitry in the JMX could cause the generation of high-level spurious tones for which appropriate shielding and filtering might prove unachievable.

3.10 Sync-distribution circuit

The sync-distribution circuit is shown in Fig. 4; amplitudes are indicated in dBm. Three reference signals at 1.024, 2.560, and 20.480 MHz are cabled to the input from the JFS. The 2.560-MHz signal is amplified, filtered, and passed through splitting hybrids to provide three −5-dBm output signals, which are then cabled to the appropriate (A or B) carrier supplies, where they are used as needed. The 20.480- and 1.024-MHz signals are amplified, filtered, and mixed in a 17A modulator to provide an output at 21.504 MHz. This signal is filtered, amplified, and passed through splitting hybrids to provide three −5-dBm output signals, which are cabled to the appropriate carrier supply inputs. Level adjustments are provided as required. Local test points are provided for the combined three-tone input signal, for both output signals, and for both intermediate signals through 30-dB bridging test pads. All these test points appear at jacks mounted in the face plate of the sync-distribution circuit drawer.

3.11 Jumbogroup 1 carrier supply

The JMX bay is arranged to process three jumbogroup signals in any arrangement; that is, it can process three jumbogroup 1 signals, one each of jumbogroups 1, 2, and 3, etc. There is no fixed assignment of the equipment to limit the flexibility with which the JMX is used. Since it is not known *a priori* which jumbogroup any given position will
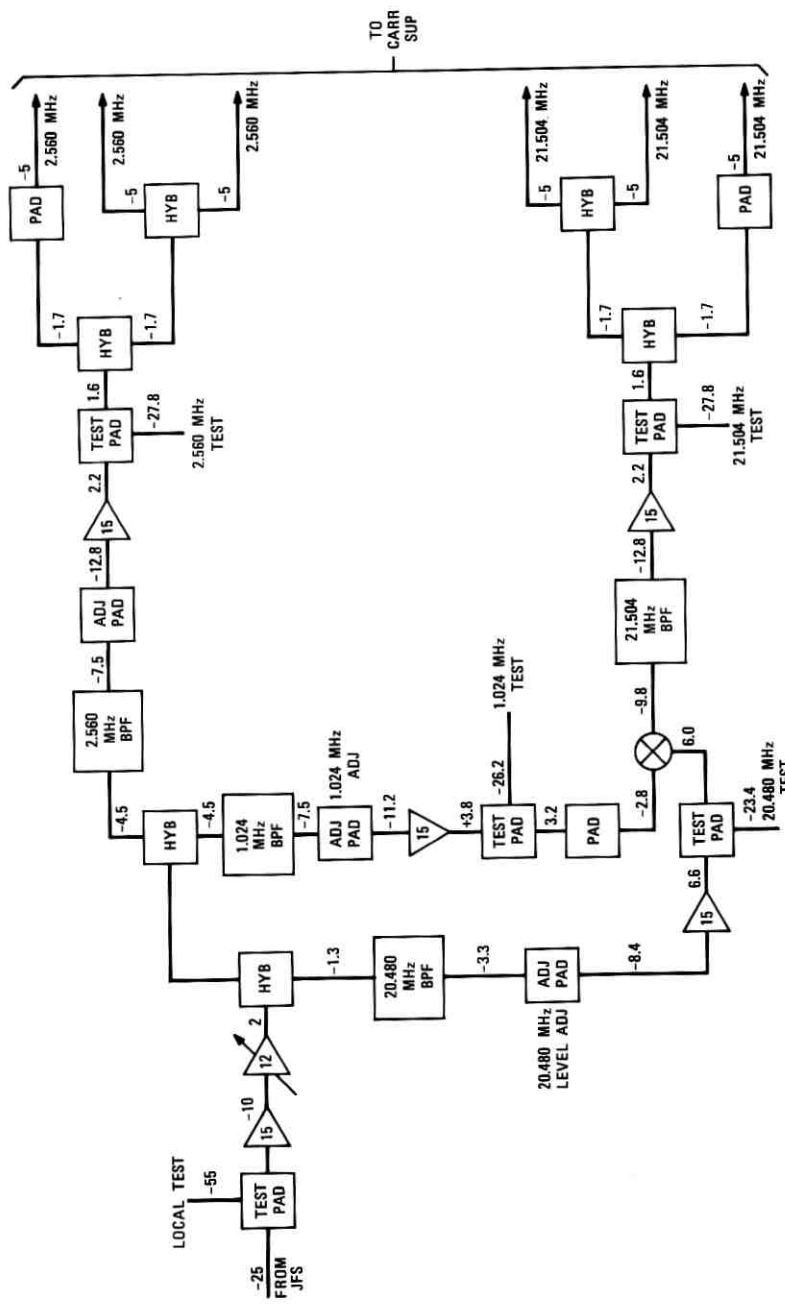
Fig. 4—Sync-distribution circuit.

handle, the sync-distribution outputs are fed to all locations. Those which are required are used. Those not required are automatically terminated when the carrier supply drawers are installed.

The jumbogroup 1 carrier supply terminates the 21.504-MHz input signal, as shown in Fig. 5; amplitudes are indicated in dBm. This carrier supply contains a temperature-compensated crystal oscillator that generates a -15-dBm, 42.496-MHz signal whose frequency stability is typically good to one part in 10^6. As with all dc-powered apparatus used in the JMX, this oscillator uses feed-through filters on the battery leads to prevent coupling between circuits through common dc paths. The oscillator output is amplified and passed through a splitting hybrid. One of the hybrid outputs passes through a fixed pad, a level adjustment, a phase-adjust network, a low-current 14-dB amplifier, a crystal bandpass filter, a 15-dB amplifier, a bridging test pad, and a splitting hybrid whose outputs each experience further noncrystal filtering to yield two 0-dBm output signals.

The phase-adjust network is provided to facilitate the matching of the A and B side signals from the modulators (demodulators) prior to manual operation of the protection switch. The match is made to minimize the mean-squared error between the output signals to achieve hitless switching. This will be covered in more detail later.

As indicated earlier, the spectral purity of the carrier signals and isolation between carrier ports are of special importance. It may appear desirable to reverse the order of the crystal filter and the 15-dB amplifier which follows it. The amplifier introduces both thermal noise and harmonic distortion. However, the signal level into the crystal filter would then be on the order of 10 dBm. The reliability of a crystal degrades significantly when it is exposed to levels exceeding 0 dBm. Accordingly, the arrangement shown was used. While the crystal filter provides sufficient rejection in the immediate vicinity of the carrier, achieving sufficient rejection elsewhere requires further filtering. This is accomplished through the use of noncrystal LC filters placed in both output legs. This output arrangement provides a net isolation between carrier supply outputs equivalent to the trans-hybrid loss of the transformer plus twice the output filter loss.

The second output of the input splitting hybrid is mixed with the 2.560-MHz input in a 17A modulator to yield a product at 45.056 MHz. This signal is similarly amplified, crystal filtered, further amplified, and passed through a 30-dB bridging test pad. The low-level output is fed to a 6-dB splitting pad that provides two test appearances, one local and one remote. The higher-level output is fed to a splitting
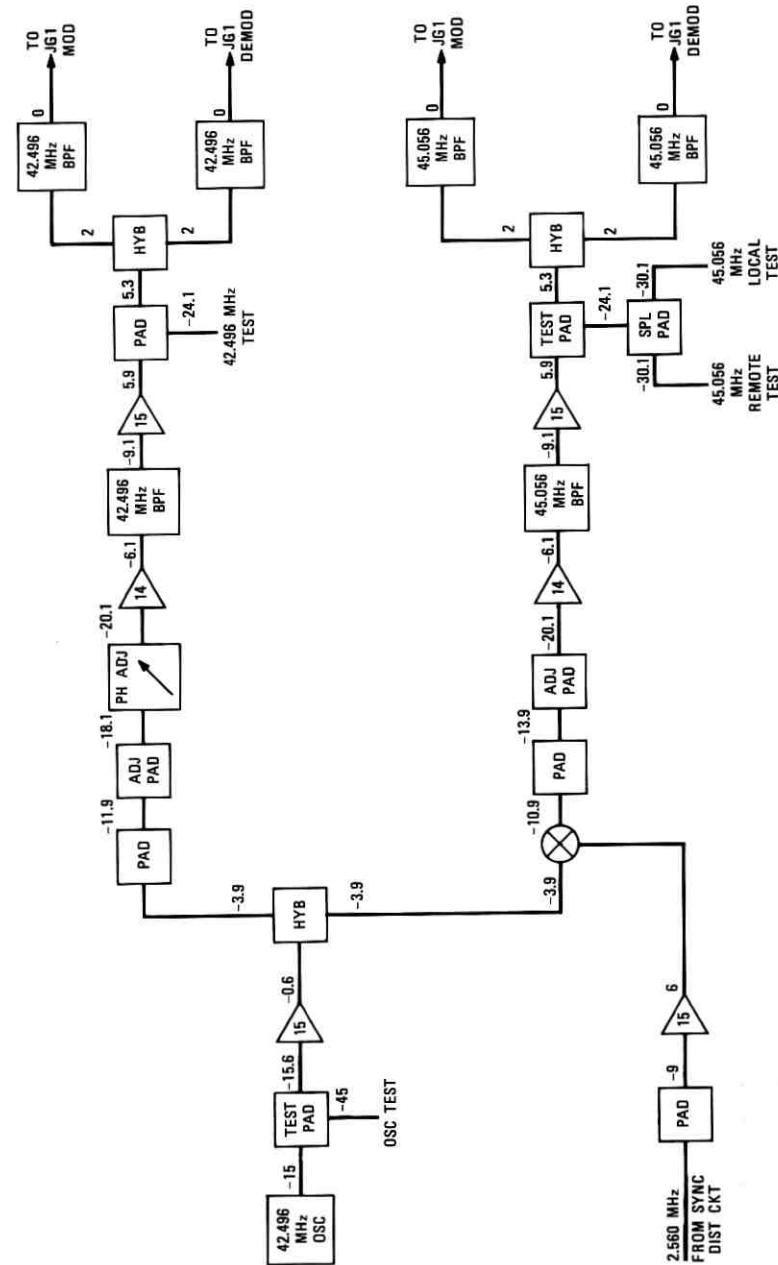
Fig. 5—Jumbogroup 1 carrier supply.

hybrid, where its outputs are further filtered to yield two 0-dBm carrier signals.

The carrier signals are cabled to the appropriate modulators and demodulators as indicated. On the transmitting side, the first modulator uses the 42.496-MHz carrier, while the second uses the 45.056-MHz carrier. Each of these carriers is only as stable as the stability of the local oscillator. However, the effective carrier frequency, which is the difference between the actual carriers used, is orders of magnitude better since it depends only on the stability of the JFS. For purposes of accurate demodulation, the stability of the local oscillator is not important. Other considerations—namely, the matching between filter rejection peaks and the location of the tones to be rejected— require that the actual carrier frequencies do not wander too far from their nominal values. The carriers must be at least as stable as a few parts in 10^6.

3.12 Jumbogroup 2 carrier supply

Functionally, the jumbogroup 2 carrier supply is very similar to that for jumbogroup 1, as shown in Fig. 6. In this case, the 2.560-MHz input is terminated and the 21.504-MHz input is mixed with the locally generated 70.144-MHz signal. For jumbogroup 2, the phase-adjust network is located in the path of the input reference signal prior to mixing with the locally generated signal, while for jumbogroup 1 the phase-adjust network was located in the through path of the locally generated signal. Functionally, either location is acceptable. The relative positions shown were chosen because they facilitated network realization.

3.13 Jumbogroup 3 carrier supply

Jumbogroup 3 requires three carriers for modulation and one carrier for demodulation; the first two steps of modulation for jumbogroup 3 are identical to those for jumbogroup 1. As shown in Fig. 7, the 42.496- and 45.056-MHz carriers are generated in the same way as for jumbogroup 1. Since only one appearance of each signal is required, no splitting hybrids are needed at the outputs. Following the second step of modulation, the jumbogroup signal has a net translation of 2.560 MHz. The third step of modulation and the one step of demodulation must both use carriers dependent solely on the JFS. The desired net translation for jumbogroup 3 is 43.008 MHz. This requires the generation of a 40.448-MHz carrier for the third step of modulation, and a 43.008-MHz carrier for demodulation. Both sync-distribution
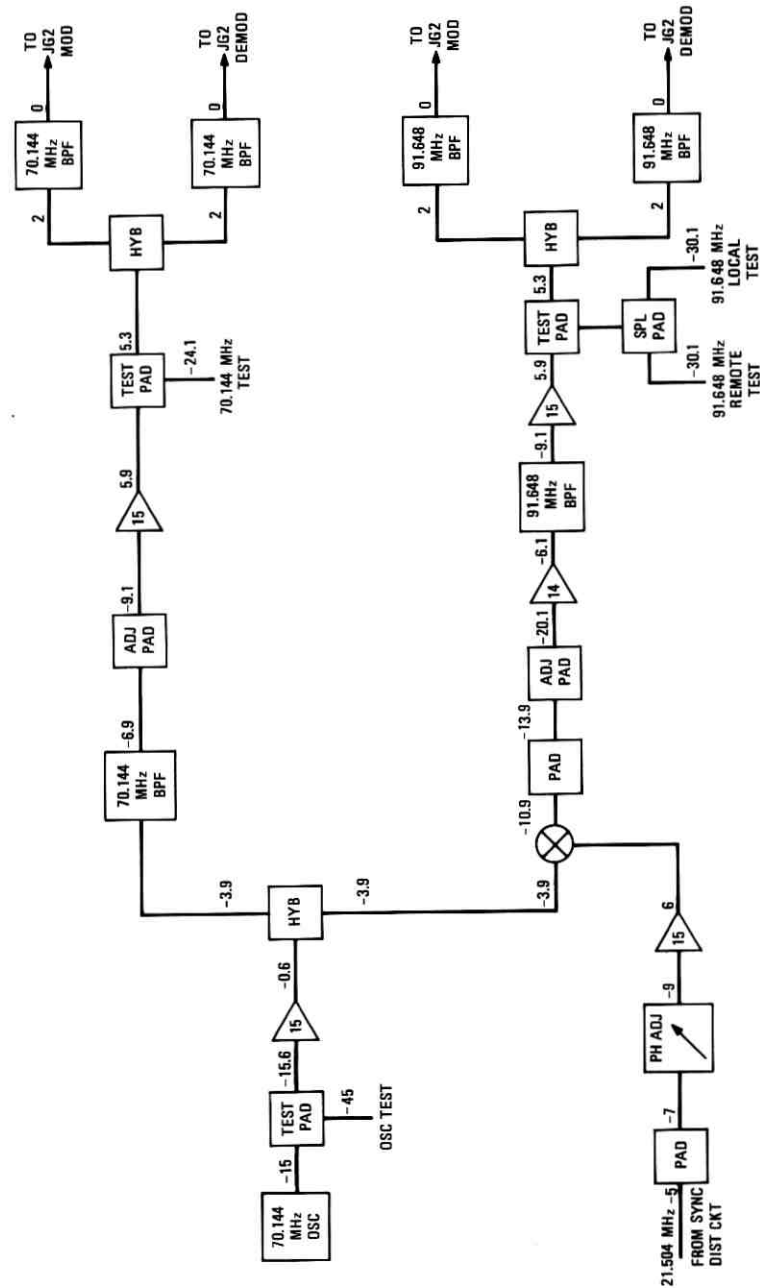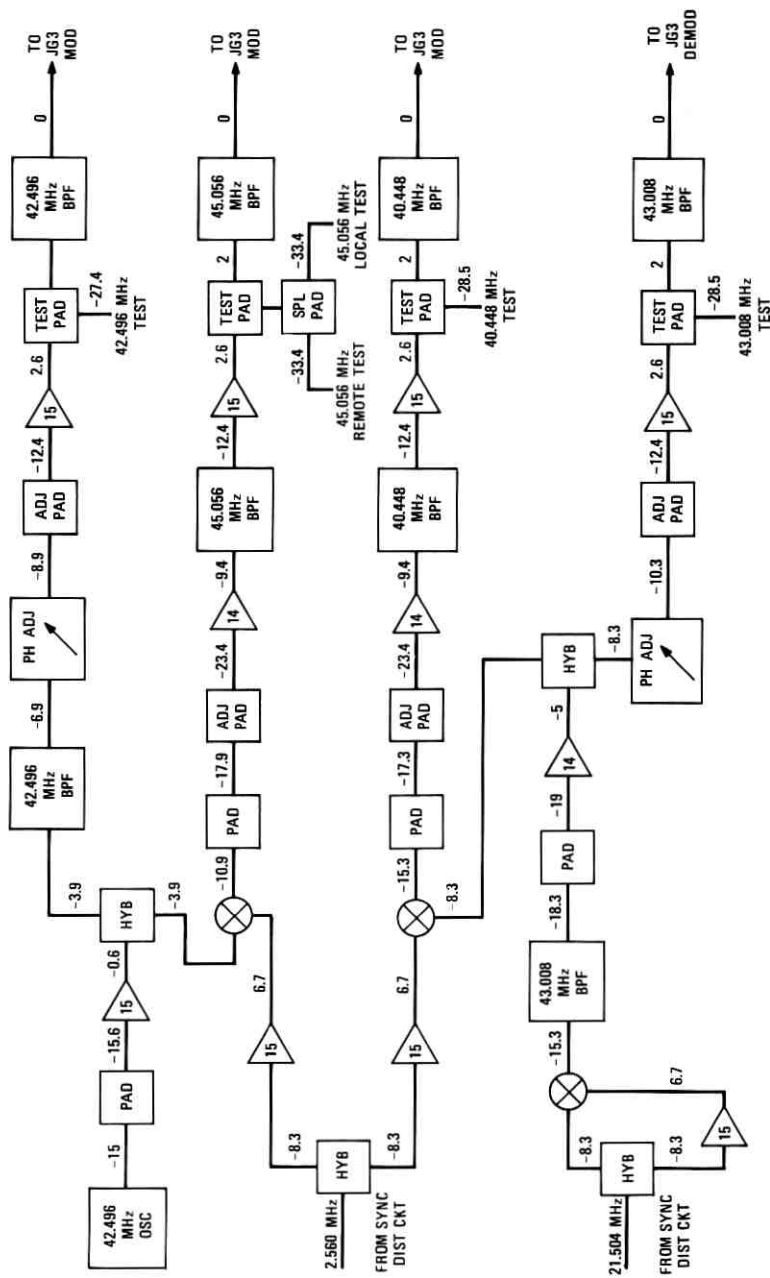
Fig. 6—Jumbogroup 2 carrier supply.
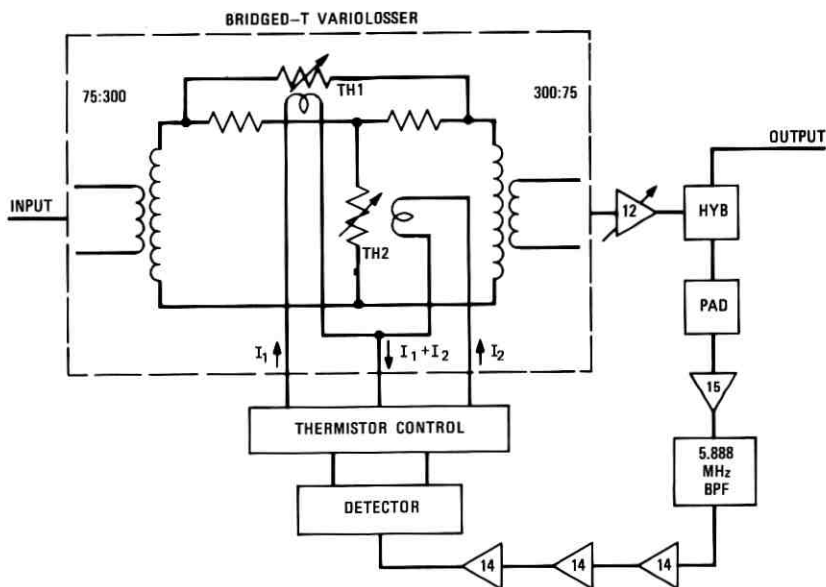
Fig. 7—Jumbogroup 3 carrier supply.

Fig. 8—Basic jumbogroup regulator.

signals are used. The 43.008-MHz carrier is obtained by doubling the 21.504-MHz input. The 43.008-MHz signal is then mixed with the 2.560-MHz input to form the 40.448-MHz carrier. Once again, all net translations depend only upon the stability of the JFS. Two phase-adjust networks are required for jumbogroup 3, one for modulation and one for demodulation.

3.14 Jumbogroup regulator

The jumbogroup regulator, shown in Fig. 8, compensates for level variations that might be introduced in the jumbogroup passband by the L5 line or terminal equipment between the remote transmitting BJGT and the regulator input. A bridged-T variolosser, using indirectly heated thermistors in the shunt and bridging arms, provides a level control of ±5 dB dynamic range. Control currents are automatically adjusted in a complementary manner in accordance with the level of the received 5.888-MHz jumbogroup pilot. In the event that the jumbogroup pilot is lost, the regulator automatically assumes a mid-range setting. An operational amplifier, connected as an integrator, provides high dc gain in the control loop. This results in a residual error of less than 0.01 dB over the full regulation range.

To facilitate maintenance, the jumbogroup regulator is equipped with a manual reset feature. Operation of the reset switch locks the regulator in the midrange state. This feature is particularly valuable when noise loading tests are being made. An indicator lamp is lighted when the regulator is manually placed in the reset state.

3.15 Protection switching

For each jumbogroup, both transmitting and receiving, A and B paths are provided. These paths emanate from a splitting hybrid and terminate on a 2 × 2 protection switch. It is the principal function of the protection-switching circuitry to monitor the integrity of the A and B paths, to assure that a good signal is fed from the protection switch when possible, and to activate alarms as appropriate. The circuitry also provides for the control of restoration switches and the manual operation of the protection switch in a hitless manner.

As shown in Fig. 9, both working and idle output jumbogroup pilot signals are monitored by the protection-switching circuitry. The switch control detector emits a dc signal proportional to the level of the jumbogroup pilot. On the transmitting side, the pilot frequencies are 8.448, 27.392, and 48.896 MHz for jumbogroups 1, 2, and 3, respectively; receiving, all jumbogroup pilots are at 5.888 MHz. If the idle signal is lost for more than 0.1 second, a minor alarm condition is indicated. If the working output is lost and the idle output is present, the switch changes state. If this causes a good signal to be present at the working output, and the signal is not present at the idle output, a minor alarm is indicated. If both outputs are lost simultaneously for more than 0.1 second, the switch does not change state, but a major alarm is indicated. Alarms are indicated both in real time and with memory. Switching is nonrevertive; that is, once a service-providing condition has been established, the switch will remain in that state; neither A nor B output is preferred. This arrangement avoids unnecessary hits on jumbogroups that might be heavily loaded with digital signals.

In the event that the working output appears lost while the idle signal is present and operation of the protection switch fails to correct the situation, no further switching occurs. Such a condition would most likely be due to failure of the circuitry monitoring the working output. A major alarm would be indicated.

Manual control of the protection switch is provided. If no trouble condition is indicated, manual operation of the appropriate control switch will cause the protection switch to change state. To facilitate
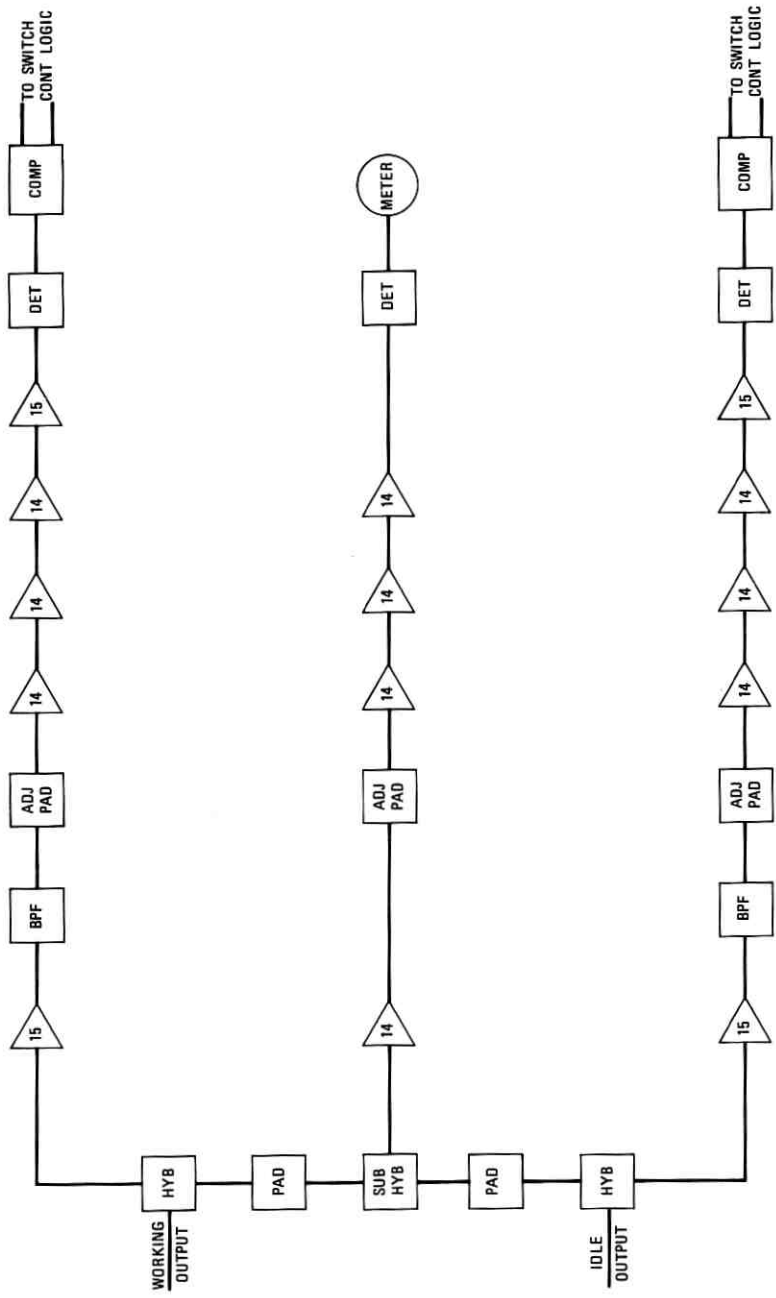
Fig. 9—Protection switching.

this operation in a hitless manner, a variable-sensitivity monitor of the degree of match is provided. The idle and working outputs are subtracted in a special hybrid transformer, yielding an error signal. This error is amplified and detected, and the resultant dc signal is fed to a meter. The meter reading is indicative of the energy in the error signal. Given that the gain of the A and B sides are closely matched, the meter reading is indicative of the phase difference between the A and B sides. Adjustment of the phase network in an associated carrier supply corresponds to varying the phase, constant with frequency, of one side relative to the other. By altering the phase network adjustment to null the meter reading, the best possible phase match is achieved. Since this procedure uses the actual message loading present at the time, it results in the best match over the most meaningful portion of the jumbogroup passband. Inability to reach a very low meter reading is indicative of gain mismatch. This can be corrected by making gain adjustments as appropriate, through the use of the bridged test jacks provided on all modules.

Once A and B paths have been matched, the protection switch is operated manually. In this mode of operation, the switch contacts are sequenced in a make-before-break manner to avoid the momentary open that might otherwise result.

If the manual operation would cause the protection switch to select an apparently failed side, the control logic will not normally respond to the manual command. In this instance, an additional, recessed button must be depressed by the craftsperson to allow him to override the inhibit feature of the control logic.

Restoration access switches are operated via the switch-control logic. Upon command for restoration switch operation, the switch-control logic locks the protection switch in its existing state. It then operates those restoration switches associated with the unused side. When the control logic is in this state, an indicator lamp is lighted at the bay.

3.16 17A modulator

Frequency translation is achieved in the JMX through the use of a double-balanced diode-ring modulator, the 17A. The important performance parameters of such a modulator include conversion loss, passband distortion, intermodulation distortion, carrier balance, and signal balance.

Conversion loss, which gives rise to the difference in level between the input signal and the desired output signal, has a theoretical mini-

mum value of 3.9 dB as a result of the partitioning of input energy into the multitude of sidebands generated by the modulation process. Additional loss is introduced principally through diode dissipation, transformer loss, and whatever additional padding is introduced owing to other considerations. While conversion loss can be offset by adding appropriate gain in the amplifiers preceding and following the modulator, the tendency of increasing conversion loss ordinarily is to result in increased noise, either because levels are increased internal to the modulator, or because the effective noise figure of the following amplifier is degraded, or both.

Passband distortion is introduced principally through transformer roll-off and circuit parasitics. Mutual coupling decreases with decreasing frequency, and shunt capacitance becomes dominant as frequency is increased. Although transformer design can be optimized for a specific frequency range, the design of a single transformer for the multitude of input and output spectra encountered in JMX modulators is particularly challenging. While passband distortion is usually equalizable, prudent design requires that such distortion be minimized at its source.

Intermodulation distortion is influenced by diode nonlinearities, diode match, transformer and circuit balance, carrier waveform, and signal level.[9] Intermodulation distortion may fall in-band or out-of-band, relative to the desired sideband. Unlike most other parameters of interest, in-band intermodulation distortion cannot be mitigated once it is generated. Consequently, this type of distortion must be controlled at its source.

Ideally, in a double-balanced modulator, neither input signal nor carrier should appear at the output. Circuit imperfections preclude realization of these objectives. Carrier balance and signal balance depend on diode match, transformer balance, and circuit layout. Typically, the carrier input level is 70 to 80 dB higher than the level of the output signal. These relative levels are chosen because they result in optimal intermodulation performance. Furthermore, JMX carriers run as high in frequency as 91.648 MHz. Consequently, sizable levels of carrier signals may appear at the output. Since they may subsequently fall within a message spectrum, such carrier leak must be attenuated. While carrier leak is theoretically filterable, rejection peaks well in excess of 100 dB are required. Furthermore, unfiltered carrier leak may significantly increase the total power that an amplifier directly following a modulator must handle. This can give
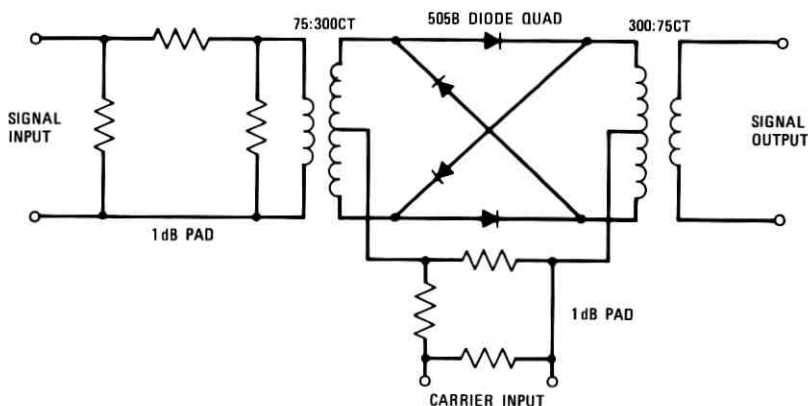
Fig. 10—17A modulator.

rise to further nonlinear distortion in the amplifier. Finally, any noise entering the carrier port from the carrier-drive amplifier will appear at the output to the extent that carrier balance allows. Should a portion of this noise overlap the output message spectrum, it will contaminate the signal in an unfilterable manner. Accordingly, good carrier balance is critical, second only in importance to intermodulation distortion.

Signal leak, by design of the modulation process (i.e., the selection of modulation steps and carrier frequencies), is always filterable. Nevertheless, signal leak at times gives rise to extremely demanding distrimination objectives.

The 17A modulator is shown in Fig. 10. The center-tapped transformers have a 75:300-ohm impedance ratio with a 7:7 + 7 turns ratio. These transformers introduce less than a 0.05-dB passband shape over any jumbogroup band. Transformer balance exceeds 55 dB.[8]

Shottky barrier diodes are used. They provide essentially instantaneous switching at all JMX carrier frequencies, and appear to be purely resistive up to 350 MHz. Their I-V characteristics are extremely linear in the region running from 2 mA to 20 mA. In this region, their resistance is about 40 ohms. Diodes are selected to yield matched quads. Each quad has less than 15 mV mismatch at 2 mA and at 20 mA.

The 1-dB pads at both signal and carrier inputs were found experimentally to result in improved intermodulation distortion, carrier and signal balance, passband distortion, and return loss.

Circuit layout was optimized experimentally. Topological symmetry was found to be critical. This was achieved, for example, by using four manually inserted crossovers as opposed to partial use of printed-circuit land connection.

The 17A modulator has a conversion loss of 6.2 dB, passband distortion not exceeding 0.1 dB, less than 0 dBrnc0 of noise at optimal signal levels, and, when driven by its associated carrier-drive amplifier, signal and carrier balance of at least 40 dB.

IV. PHYSICAL ARRANGEMENT

The JMX is housed in a shop-wired, shop-tested, 11-foot 6-inch, unitized double bay. It is arranged to contain all the equipment to process three jumbogroup signals. Each jumbogroup position is independent of other positions and contains transmitting and receiving jack fields, transmitting and receiving switch-control logic units, and redundant modulator and demodulator subsystems, carrier supplies, regulators, equalizers, transmitting and receiving line interface units, and detectors. Common equipment includes redundant sync-distribution circuits at the lower left of the bay, and a summary alarm circuit at the upper right. Sixteen dc-to-dc converters are located at the top of the bay, just above the fuse panels. Alarm cutoff and lamp buttons are at the lower right just above the writing shelf. Redundant battery feeds and JFS inputs are provided. Several photographs of the JMX equipment are contained in Ref. 11 of this issue.

V. OPERATIONAL FEATURES

No at-the-bay routine maintenance is planned for the JMX. Rather, each jumbogroup is automatically and remotely monitored by the transmission surveillance system[12] associated with L5. Eleven critical test points are provided for each jumbogroup position. These include the transmitting input, the transmitting A and B outputs, the receiving A and B inputs and outputs, the A and B regulator inputs, and two carrier supply test points. These eleven access points are concentrated in a remotely controllable 1 × 12 switch. Through this arrangement, detailed analysis of JMX performance is possible with little intervention by a craftsperson.

All the above-mentioned test points, in addition to others, are also provided at the bay. It is possible to verify the integrity of any transmission-related module without removing it from the bay, through the use of the test points provided.

Verification of the integrity of the many status and alarm indicators is facilitated through the lamp-test feature. Depressing the three lamp-test buttons allows the location of any failed lamp in seconds.

VI. JMX BAY EVALUATION

The JMX has undergone extensive field evaluation. Both functionally and operationally, the equipment operated superbly. The poorest jumbogroup noise performance obtained was 12 dBrnc0.

The poorest equal-level coupling loss obtained was 90 dB. This was measured between A and B sides of the same transmitting or receiving jumbogroup, and was due to coupling through the subtracting hybrid transformer of the circuitry used to evaluate the degree of match between the associated A and B sides. This is a well-controlled coupling, not expected to vary from system to system. Most other forms of coupling were so low as to be unmeasurable.

Extensive testing of regulation and protection switching showed these features to operate without fault.

Ten JMX bays were constructed by Western Electric for use in the Lillyville, Pennsylvania-Hillsboro, Missouri route. This equipment has been providing commercial service since January 1, 1974.

VII. ACKNOWLEDGMENT

Taking into consideration the bandwidth of the signal it processes and the frequencies at which it operates, the JMX is the quietest analog multiplex terminal ever designed. The success of the design is a tribute to the technical excellence and total dedication of its design team which included J. S. Young, T. B. Merrick, G. W. Kattke, B. B. Garg, S. J. Davis, A. G. Favale, and W. D. Radwill.

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L5 SYSTEM:

39A Precision Oscillator

By A. F. FLINT and H. S. PUSTARFI

(Manuscript received May 2, 1974)

The 39A precision oscillator is designed to operate at 40 watts from a 24-V ±10-percent power source. The reference frequency can be set digitally to 5.12 MHz ±5 × 10⁻¹¹ over a tuning range of ±4 × 10⁻⁷. The output signal is a 1-mW sinewave into a 100-ohm load. This performance is assured over a temperature range of 0 to 60 degrees Celsius. Oscillator outline dimensions for length, width, and height are 35.3, 22.1, and 21.8 cm respectively.

I. INTRODUCTION

The 39A oscillator was developed to meet the precise frequency-control requirements of the L5 jumbogroup frequency supply. The need for frequency control as exemplified by the 39A oscillator is explained in the paper on the jumbogroup frequency supply in this issue. This 5.12-MHz precision oscillator is characterized by long-term stability of $<1 \times 10^{-10}$ per day, and by short-term stability of $<1 \times 10^{-8}$ for a 1-millisecond sampling time or $<2 \times 10^{-11}$ for a 10-second sampling time. External frequency control is accomplished utilizing a compatible digital word applied to the oscillator's 14-bit, TTL, precision, digital-to-analog converter. Setting range of the converter is $> \pm 4 \times 10^{-7}$, which gives a least-significant-bit resolution of $\approx 5 \times 10^{-11}$; the converter section incorporates a linearizing network to reduce the frequency-set deviation from a straight-line function to $< \pm 5 \times 10^{-11}$ over a $\pm 4 \times 10^{-8}$ tuning range, and to a maximum variation of tuning sensitivity of ± 30 percent over the entire range of $\pm 4 \times 10^{-7}$, with less than $\pm 1 \times 10^{-10}$ departure from a smooth curve.
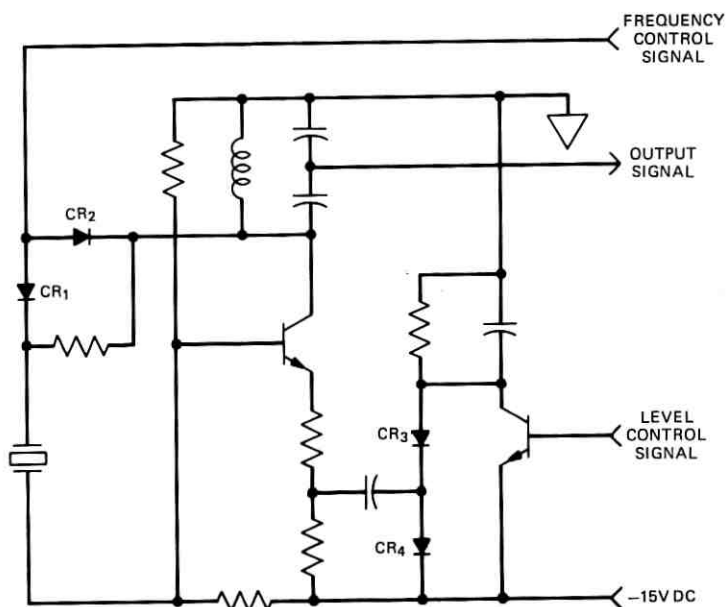
Fig. 1—Oscillator schematic.

II. GENERAL DESCRIPTION

The 39A oscillator can be separated into three major subassemblies:

(i) An oscillator RF subassembly consisting of a crystal oscillator circuit, a tuned feedback amplifier detector circuit (AGC) functioning to provide constant low-level crystal-current drive, and an output amplifier instituting sufficient load isolation for the output signal.

(ii) A thermal and electrical noise-isolation subassembly which is composed of temperature-control circuits, power-supply regulators, and redundant electrostatic shielding amply protecting all major circuitry.

(iii) The digital frequency-control subassembly with a precision digital-to-analog converter and a frequency-linearizing network for remote steering of output frequency.

III. OSCILLATOR SUBASSEMBLY

The crystal-oscillator circuit in the 39A oscillator is the modified Pierce commonly used in precision oscillators for many years.[1,2] Figure 1 is a functional schematic of the oscillator-circuit board. A

varactor diode in the crystal network is used to control the operating frequency of the network. Earlier circuit designs have varied the oscillator-transistor dc operating point to control the loop gain, but here low-noise performance is achieved by fixing the dc operating point of the oscillator transistor and adjusting its positive loop gain. Oscillator loop gain is controlled by varying the forward bias of variolossers CR_3 and CR_4, which controls the degeneration of this transistor stage.

The AGC amplifier circuit is shown in Fig. 2. This is a three-stage direct-coupled feedback amplifier selected for low noise, simplified tuning, and having a minimum number of components. Passband tuning is accomplished by an LC network in the feedback path. This amplifier operates with 38-dB in-band gain and has a 3-dB bandwidth of about 200 kHz. The rectified output from diodes CR_1 and CR_2 is used to control the oscillator gain for a crystal-operating current of 100 to 200 μA.

The output signal of the AGC amplifier is coupled to the final amplifier circuit by a variable attenuator which is used to adjust the oscillator output level to 1 mW into a 100-ohm load. The design of the output amplifier is essentially the same as the AGC amplifier.

The crystal oscillator, AGC amplifier, and output amplifier combine to make up the oscillator subassembly shown in Fig. 3. The unit at
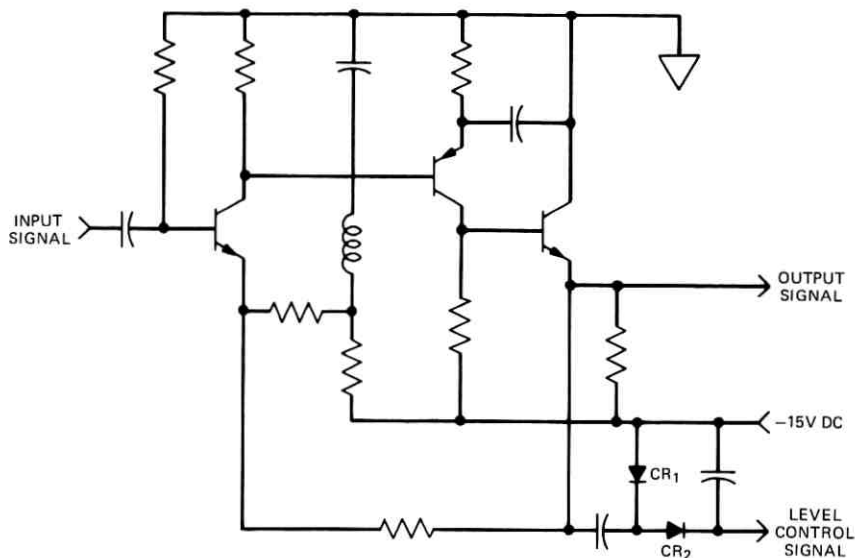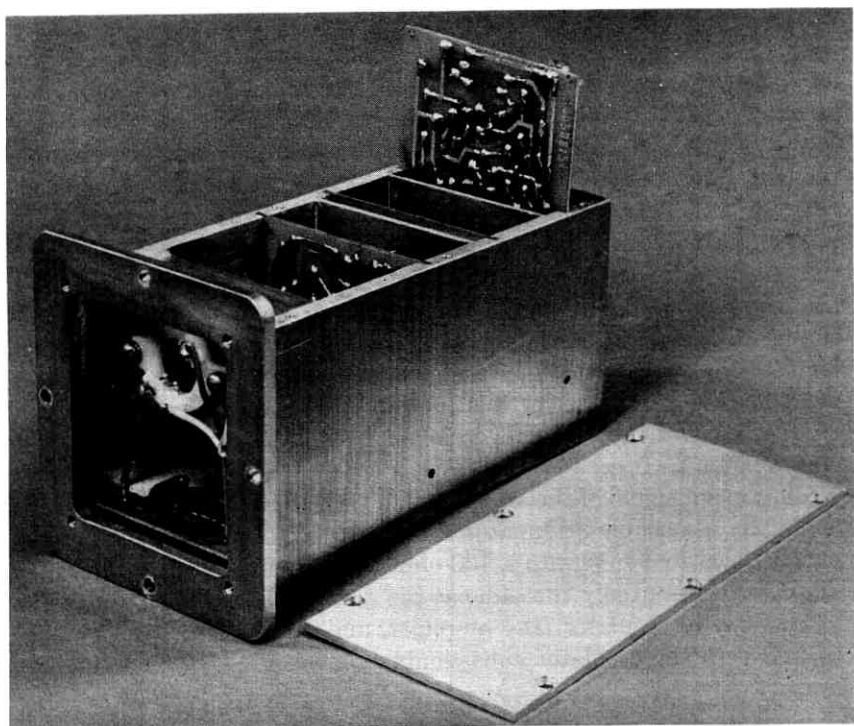

Fig. 2—AGC amplifier circuit.

Fig. 3—Oscillator RF subassembly.

this stage in fabrication can be production tested as a functional subassembly. These tests include a thorough mechanical inspection, a test of AGC and output-level, and a check on harmonic distortion.

The long-term frequency of this precision oscillator is determined primarily by changes of the crystal unit and directly associated LC components. In this respect, these components have been selected for minimum long-term drift. Short-term frequency variations (times of less than 0.1 second) can be attributed to transistor noise and small fluctuations in crystal current. The effects of crystal-current variations have been reported[3] and are typically 1×10^{-11} per 0.02 dB for current levels in the range of 100 to 200 μA at a frequency of 5.12 MHz.

IV. THERMAL AND ELECTRICAL NOISE-ISOLATION SUBASSEMBLY

The 5.12-MHz crystal units[4,5] used in the 39A oscillator operate at 76°C minimum to 81°C. At this operating temperature, the frequency dependence on temperature is approximately $7 \times 10^{-9}/$°C if the operat-

ing temperature of the crystal is $< \pm 0.25°C$ from the frequency turning point. In addition to the effects of long-term temperature variations, short-term changes in temperature cause perturbations in frequency due to thermal gradients within the crystal. A 1×10^{-11} frequency change results from a $4 \times 10^{-3}°C$/hour rate of change in crystal temperature.

To reduce frequency variation effects due to ambient temperature changes, the oscillator subassembly is housed in a two-stage temperature-stabilized oven. This two-stage oven has an ambient ratio of $> 10^5 : 1$ resulting in short-term temperature variations being measured as $< 1 \times 10^{-4}°C$ per hour. The outer oven reduces ambient effects on the temperature-control and voltage-regulator circuits to $< 5 \times 10^{-2}°C$. Two identical voltage regulators are used in the oscillator to reduce

ENVIRONMENT ISOLATION SUBASSEMBLY

OUTER OVEN

18V REG

| OUTER OVEN TC | INNER OVEN TC | VOLTAGE REG 15V REG | VOLTAGE REG 18V REG | LOW–PASS FILTER | 24V |

LOW–PASS FILTER   LOW–PASS FILTER

LOW–PASS FILTER   D/A NETWORK

OSC   AMP   AMP   BAND–PASS FILTER f_o

AGC
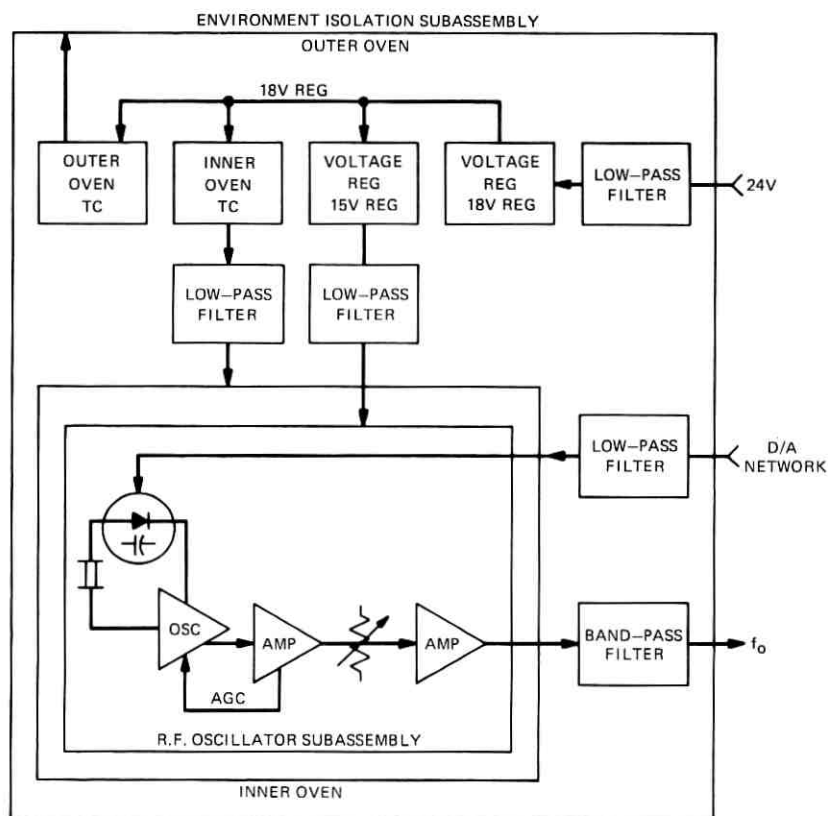
R.F. OSCILLATOR SUBASSEMBLY

INNER OVEN

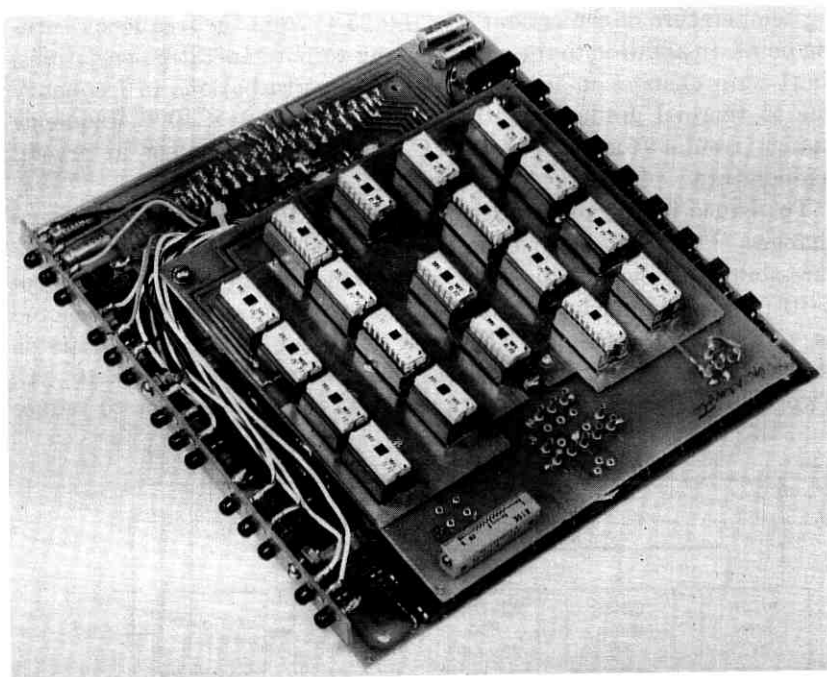Fig. 4—Precision oscillator functional components.

Fig. 5—Digital frequency-control subassembly.

the effects of power-supply variation on reference voltages, RF circuits, and control circuits. Figure 4 shows the environmental isolation subassembly.

V. DIGITAL FREQUENCY-CONTROL SUBASSEMBLY

Frequency control is accomplished by a varactor diode in the crystal circuit. The precision voltage supplying this diode is derived from the precision digital frequency-control subassembly shown in Fig. 5. This subassembly consists of a 14-bit digital gate network, with strobing option, fabricated from standard TTL logic. The network contains latching circuits, relay drivers, and relays (5 V dc) controlling a constant-impedance voltage divider. Figure 6 gives the basic circuit diagram for this digital-to-analog converter.

Figure 7 is a simplified schematic of the divider excluding the control logic. Resistors R_{V1}, R_{V2}, and R_{M1} (or R_{M2} if required) are selected by computer program at time of manufacture to precisely give the required nonlinear binary-voltage allowing a linear binary-frequency function of oscillator output. The loading resistor, R_{M1} or R_{M2}, de-
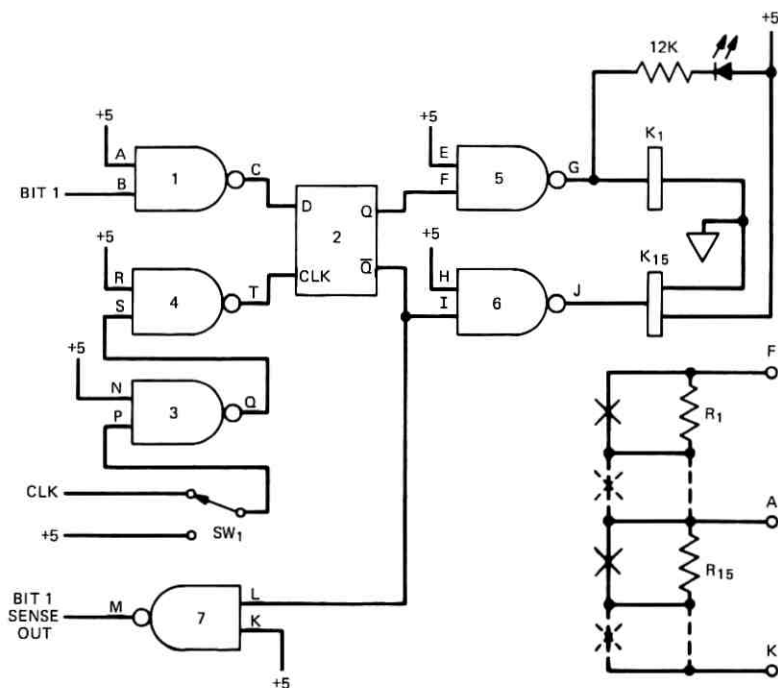
Fig. 6—39A oscillator simplified circuit.

termines the network output voltage function and is selected for each oscillator. A typical set of output-voltage characteristics for this network are shown in Fig. 8 for values of R_{M1} or R_{M2} from 20,000 ohms to ∞ plotted as output voltage versus ratio of the programmed binary divider.

Figure 9 shows the tuning characteristics of a typical 39A oscillator over the frequency range from $F_o (1 - 4 \times 10^{-8})$ to $F_o (1 + 4 \times 10^{-8})$ and plotted relative to the frequency setting. The mean slope achieved over this range is within a few percent of the objective, with slope uniformity within about ± 10 percent over the range. Over the entire
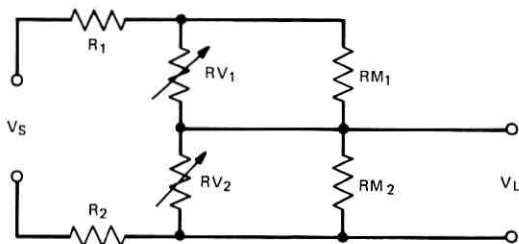

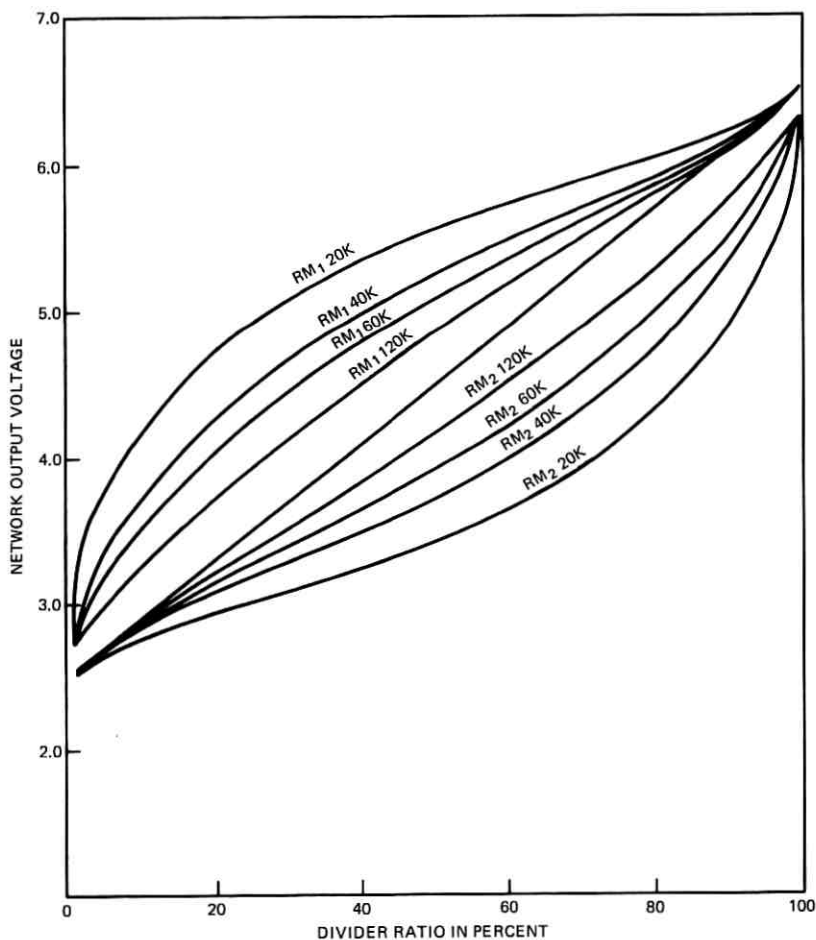
Fig. 7—Simplified divider schematic.

Fig. 8—Typical digital-to-analog converter output-voltage characteristic.

tuning range of $\pm 4 \times 10^{-7}$, slope variations are no more than ± 30 percent, and deviation from a smooth curve is no greater than $\pm 1 \times 10^{-10}$.

The circuit also provides an LED indicator for each input bit, which allows a visual check of the frequency-setting word. This, in addition to an electrical output provision for automatically sensing the tuning status, gives a convenient check on the complete functioning of the digital-to-analog converter.

VI. ELECTROMECHANICAL DESIGN

The overall electromechanical design makes maximum use of current instrument-fabrication methods. All printed-circuit boards are of a plug-in configuration simplifying assembly and testing by a large factor. Hand wiring has been held to a minimum throughout the entire assembly. Polyurethane foam both isolates and supports the inner and outer oven assemblies. RFI or constant K bandpass filtering has been incorporated between major subassemblies where deemed necessary.

Figure 10 illustrates the entire oscillator disassembled to show inner and outer ovens plus the inner and outer card frames. The front cover is mechanically attached to the digital-to-analog converter. The two connectors seen on the precision digital-to-analog converter circuit-board serve to provide a major part of the wiring interconnecting the rear input panel and the oscillator circuitry. The right front vertical edge of the external housing is designed to allow viewing of the LED
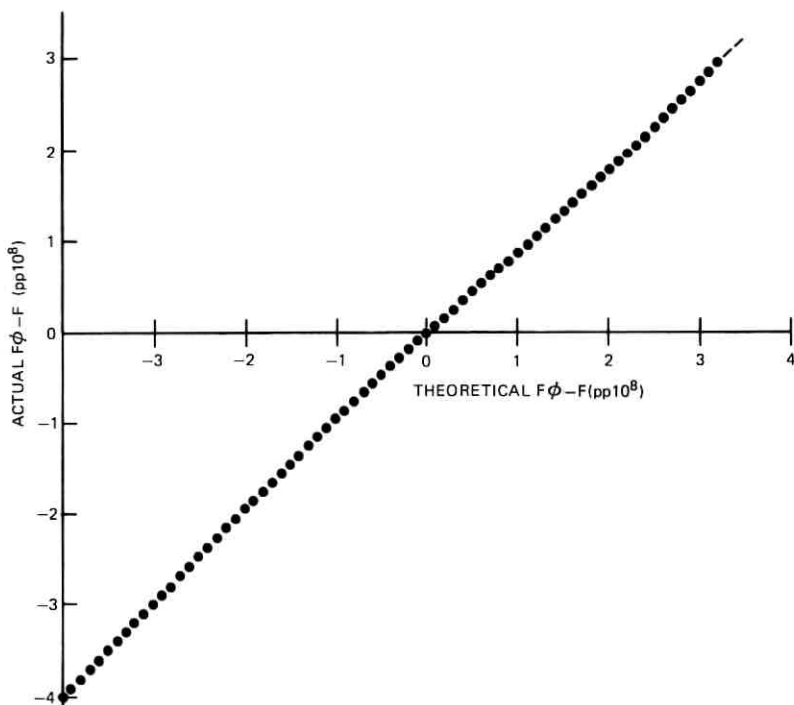

Fig. 9—Measured frequency range vs theoretical frequency range.
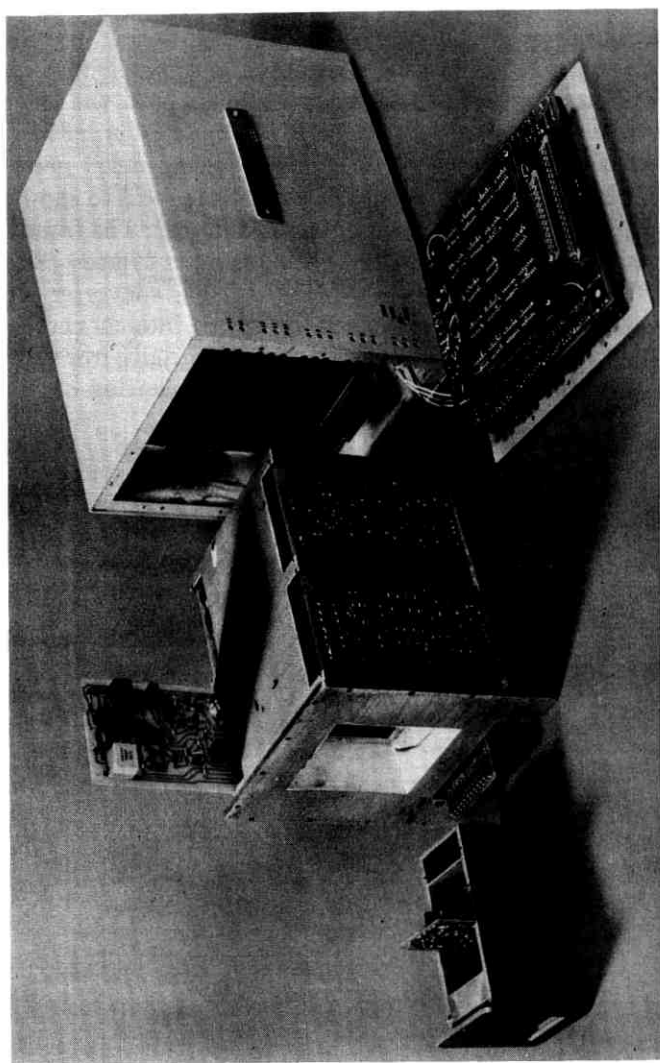
Fig. 10—Disassembled 39A oscillator.

indicator. All connections are made through the rear oscillator panel to the digital-word inputs, strobe pulses, 5- and 24-V dc power inputs and to the sense and RF outputs.

VII. SUMMARY

The 39A oscillator has been developed as a secondary frequency standard and is characterized by long-term stability of better than 1×10^{-10}/day, and short-term stability of better than 1×10^{-8} for a 1-millisecond sampling time and 2×10^{-11} for a 10-second sampling time. The total range of digital-to-analog frequency control is $\pm 4 \times 10^{-7}$ from a nominal 5.12 MHz having slope variations $< \pm 30$ percent with a maximum deviation of any point from a smooth curve of less than 1×10^{-10}. Midrange frequency control has a nominal straight-line tuning characteristic of 5×10^{-11}/bit, with an average slope tolerance of ± 10 percent and maximum deviation for a smooth curve of less than $\pm 5 \times 10^{-11}$. This oscillator has been designed to minimize the effects of thermal, mechanical, and electrical noise to a high degree and is fabricated using quality hardware and reliable components.

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L5 SYSTEM:

Jumbogroup Frequency Supply

By J. F. BARRY, S. NARAYANAN, and J. F. OBERST

The jumbogroup frequency supply (JFS) provides accurate and reliable synchronization signals to L5 jumbogroup multiplex carrier supplies. The JFS output-signal frequency is held to within several parts in 10^{10} of an input reference frequency. This extraordinary accuracy is achieved by digital frequency control of a precision quartz oscillator. The excellent reliability required when the JFS serves over 100,000 two-way voice circuits is achieved by redundant power and signal feeds, one-for-one protection of key circuits, and automatic protection switching.

In addition to its L5 application, the JFS will also serve as a regional frequency supply in a new synchronizing frequency distribution network that will upgrade the entire Bell System transmission plant.

I. INTRODUCTION

It is the goal of the Bell System to maintain frequency offset in any transcontinental circuit to within ± 2 Hz. This requirement is imposed by the most sensitive message signals, such as those in programming, telegraph signals, and data signals. In an analog amplitude-modulated system, such a frequency offset can be introduced into the message signal if the transmitting and receiving multiplex carriers are not synchronized. In a single-sideband suppressed carrier multiplex format as used in jumbogroup multiplex[1] terminals (JMX), the multiplexing carriers are not transmitted. Thus, precision frequencies must be provided for the multiplex terminal at each geographical location.

The top channel frequency of the long-haul, analog, frequency-division multiplex system has been steadily increasing. With the increase in top channel frequency (F), the frequency precision required of the multiplex carriers increases. This is because the frequency offset

(ΔF) is limited to 2 Hz, whereas the ratio $\Delta F/F$ decreases as F increases. In the L5 coaxial system, the top channel frequency is at 60 MHz, and the 2-Hz requirement results in a ratio of 3.3 parts in 10^8 ($3.3/10^8$). Because a typical signal may traverse through several JMX terminals, carrier frequencies must be synchronized with an accuracy of about 1 part in 10^8.

The frequency precision demanded by the L5 system could not be provided by the synchronizing network existing at the time L5 design work was initiated. Several alternatives were studied. One alternative was to provide a cesium frequency standard at each multiplex location. The cost of the cesium standard, the need for providing at least two such standards for reliability at each location, and the need for training personnel to maintain them made this option unattractive. An alternative was to provide a primary frequency standard at a central location and transmit reference-frequency signals to different locations. Such a primary frequency standard, called the Bell System Reference Frequency Standard[2] (BSRFS), has been designed. It comprises three cesium standards and provides highly accurate reference signals (better than one part in 10^{11}).

At the receiving location, several alternatives were open. One option was to use a precision voltage-controlled crystal oscillator, whose frequency drift could be manually corrected periodically. The manual control implied trained personnel and periodic maintenance. Moreover, frequency-control circuitry is required for alarming and switching purposes in case of oscillator failure. Thus, it was decided to control and monitor frequency automatically, and the precision frequency supply so designed (the jumbogroup frequency supply, or JFS) is the subject of this paper.

The reference-frequency signal, even though highly accurate, may suffer short term impairment like noise, switching transients, etc. It may even be lost completely due to transmission difficulties. Because of the large number of circuits (more than 100,000) dependent upon one JFS, it is imperative that precision frequencies be supplied to the JMX even when the reference signal is lost or impaired. To accomplish this objective, a precision crystal oscillator capable of accurately free-running for a short period of time is employed in the JFS. The precision oscillator is loosely coupled to the reference signal such that it can automatically correct itself when the unimpaired reference signal is present and disconnect from the reference signal if the reference signal is impaired or lost. A digital memory is provided in the digital frequency-control circuit so that the oscillator can remain at the correct
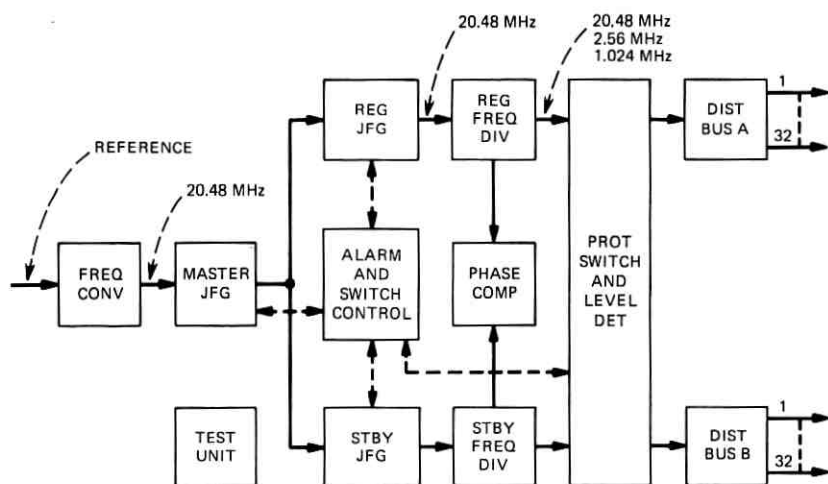
Fig. 1—Block diagram of jumbogroup frequency supply.

frequency even when the reference signal is disconnected. The precision oscillator and the precision digital frequency-control circuits are housed in a unit called the jumbogroup frequency generator (JFG).

An important consideration in the design of the JFS is reliability. More than 100,000 two-way voice circuits may depend upon one JFS. To ensure reliability, automatic protection switching is provided for its key circuits. Several visual alarms are provided so that preventive action can be taken before protection switching is needed. The working of the frequency-control circuitry can be checked without any interruption of service. Redundant power feeds connect each JFS to the office battery and two cables are provided to each JMX.

Human engineering has been given careful consideration. A precision test unit to measure accurately the difference between two signals has been incorporated in the JFS. This will enable the craftsperson to troubleshoot the bay if necessary. A manual patching arrangement and manual override have been provided to ensure that the craftsperson can take full control of the bay if needed.

II. SYSTEM DESCRIPTION OF JFS

The JFS furnishes three precision frequencies to the JMX carrier supply, where further processing occurs to generate all the carriers needed for frequency multiplexing. The three frequencies are at 20.48 MHz, 2.56 MHz, and 1.024 MHz. An overall block diagram of the JFS is shown in Fig. 1. It consists of three JFG's arranged in a master-

slave configuration, two frequency dividers, switch and alarm circuitry, two output buses, and a precision frequency-measuring test unit.

The master JFG directly receives the 20.48-MHz reference signal. In some locations, the reference signal will be at 2.048 MHz; a frequency converter has been designed to convert 2.048 MHz to 20.48 MHz. The master JFG compares its precision-oscillator frequency with the incoming reference signal and corrects its oscillator based on the long-term average of the frequency difference. The master JFG output signal is at 20.48 MHz and is fed to the regular and standby JFG's.

The incorporation of three JFG's in a JFS ensures detection of a JFG failure even when the reference signal is noisy or absent. Such a detection would quickly initiate alarm and switching circuits to take appropriate action. If only two JFG's were provided in a JFS, the failed JFG could not be isolated if the failure occurred when the reference signal was absent or noisy.

The regular and standby JFG's use the output of the master JFG as their reference. In this master-slave arrangement, the input signal to the regular and standby JFG's should be good all the time unless the master JFG fails. In contrast, the master JFG reference signal, while highly accurate when present, may be absent or noisy at times. For this reason the master JFG does not produce frequency alarms unless the frequency offset exceeds $1/10^8$ and lasts at least two hours; the regular and standby JFG's produce frequency alarms if the offset exceeds $2/10^9$ and lasts for several hundred milliseconds (see the JFG description in Section III).

As shown in Fig. 1, the output of the regular JFG (20.48 MHz) is fed to the regular frequency divider; similarly, the standby JFG feeds the standby frequency divider. The function of the frequency divider is to generate 2.56-MHz and 1.024-MHz signals and combine all three frequencies (20.48 MHz, 2.56 MHz, and 1.024 MHz) together for transmission on one coaxial cable. The output of the regular divider is transmitted to the two output buses through two coaxial switches. The coaxial switches are operated by the alarm-and-switch-control unit which processes information from the three JFG's and the level detectors. The output bus has 32 taps, and because of reliability considerations one JFS may serve up to 22 JMX's. Each JMX is connected by two cables, one from bus A and the other from bus B. Thus, the regular divider feeds both sides of the JMX. The output of the standby divider is fed to the two coaxial switches. If a regular JFG or frequency divider fails, the alarm and switch control unit will operate the coaxial switches so that the standby JFG and divider will supply signals to
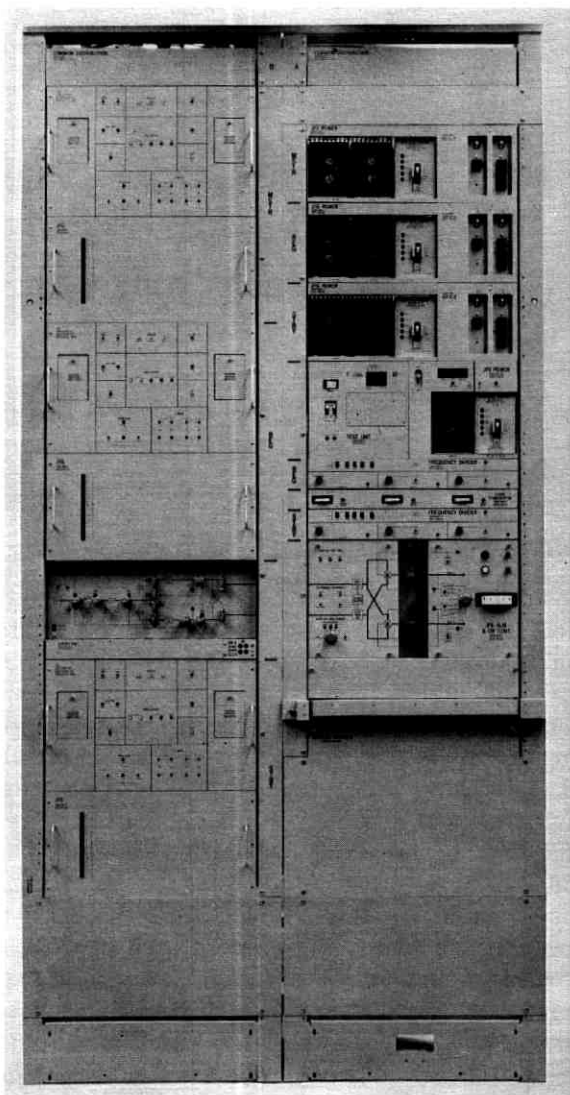
Fig. 2—Jumbogroup frequency supply bay.

both sides of the JMX. A phase comparator is provided to compare the phases of signals from the regular and standby dividers; this enables the craftsperson to manually transfer output signals from the regular JFG to the standby JFG with minimal disturbance.
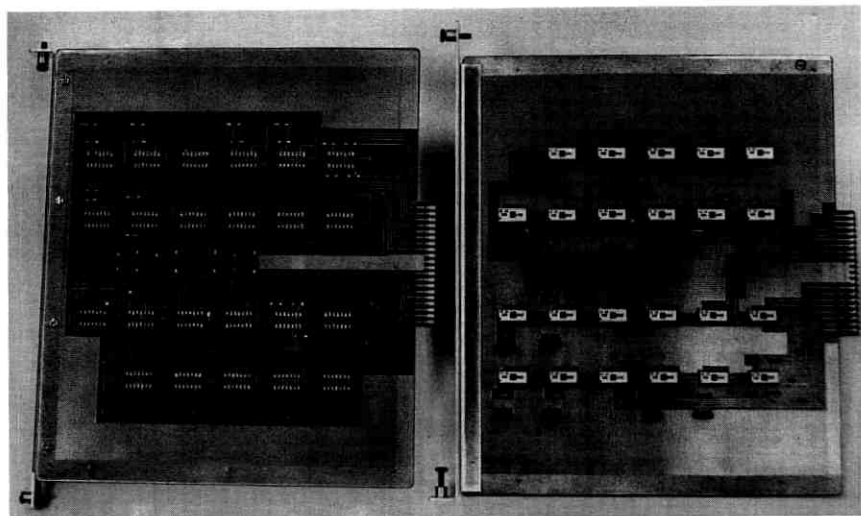
Fig. 3—Typical printed-wiring board for jumbogroup frequency supply.

Several features have been included in the JFS to facilitate trouble-shooting and maintenance. A precision frequency-measuring test unit has been designed and incorporated as a part of the JFS. A patch panel has been included to provide a convenient and non-service-affecting way to rearrange the remaining JFG's if a JFG fails. Several test points and a level meter have been provided. The coaxial switches can be patched around without affecting service; the alarm and switch control circuit could then be exercised for routine checkup of the coaxial switches and control circuitry.

The JFS is shown in Fig. 2 in a double 9-foot bay. A 7-foot-bay arrangement is also available. At the top of each side is a distribution bus; these buses provide 32 redundant (64 total) outputs. The three JFG's are on the left side with the master on top and below it the regular and standby JFG's with the JFG patch panel between them. Each JFG has its alarm display panel above its oscillator panel on which the 14 bits of the oscillator frequency-control word are displayed. On the right side, from top to bottom, are shown: three sets of dc-dc converters, test unit and associated dc-dc converters, regular frequency divider, phase comparator, standby frequency divider, and alarm and switch control unit.

Most of the JFS circuitry consists of digital integrated circuits (IC's) and various discrete components mounted on two- or four-layer

printed-wiring boards. Additional apparatus such as thin-film, hybrid, integrated-circuit amplifiers and crystal filters are also employed.

The wide use of digital IC's in the JFS made it possible to achieve sophisticated control and alarm functions that would be impractical with discrete components. Western Electric-manufactured 101-type emitter-coupled logic (ECL) was chosen for reasons of speed and availability. Since only small-scale integration (SSI) members of this logic family were available during the JFS design phase, a large number of "chips" was necessary to implement all JFS functions—about 1100 logic chips in a full JFS. A typical two-sided printed-wiring board is shown in Fig. 3.

The JFG is discussed in detail in the next section. The frequency divider, protection switching, and test unit are discussed in the succeeding sections.

III. JUMBOGROUP FREQUENCY GENERATOR

3.1 General description

The JFG produces an accurate 20.48-MHz sinusoidal output signal with a $\Delta F/F$ normally less than $1/10^{10}$ relative to its input signal. This small $\Delta F/F$ is achieved by combining an extremely stable crystal oscillator with precision frequency-control circuitry. Frequency accuracy is provided by reference to a signal whose ultimate source is the BSRFS.

The oscillator is the 39A digitally controlled crystal oscillator.[3] It has a drift rate less than $1/10^{10}$ per day, low enough to operate without correction for several weeks and still meet JFS frequency-accuracy requirements. This oscillator's frequency is controlled by a 14-parallel-bit binary-control word developed by the JFG frequency-control circuit. The control word is changed when comparison of the JFG input and output signals indicates a need to correct the oscillator frequency.

Figure 4 is a block diagram of the JFG showing the relationship between oscillator, input, output, and frequency-control circuits. The 39A digitally controlled crystal oscillator (DCXO) has a 5.12-MHz output which is multiplied to 20.48 MHz in the JFG output circuit. The 20.48-MHz input signal, after being filtered and amplified in the JFG input circuit, is compared with the 20.48-MHz output signal in the frequency-control-and-alarm circuit, which produces the frequency-control word and frequency alarms. The logic-interface circuit processes frequency, level, and voltage alarms generated within the JFG for transmission to the JFS alarm-and-switch-control circuit and for dis-
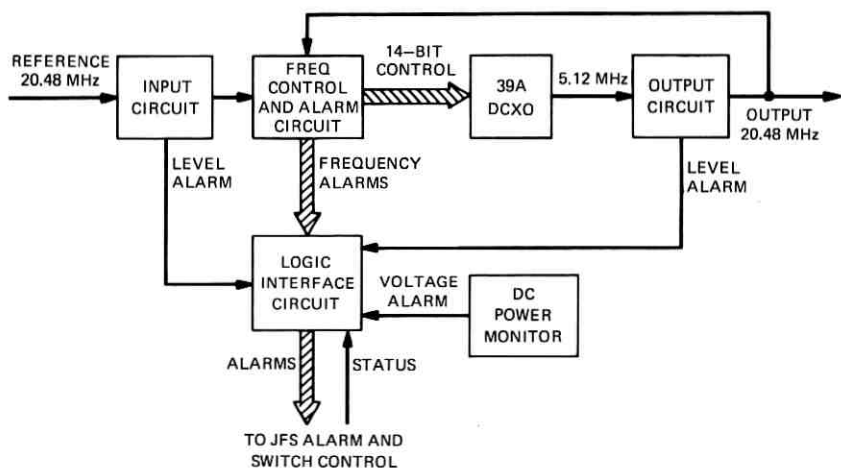
Fig. 4—Jumbogroup frequency generator.

play on the JFG itself. This alarm processing is controlled by a status input from the JFS alarm and switch control.

These frequency-control and alarm features will be described in some detail in the following sections, along with a number of other features designed to facilitate the operation and maintenance of the JFG.

3.2 Frequency control

The major consideration in designing the JFG was the character of the signal supplied to the JFS as a frequency reference. As mentioned earlier, this signal could be intermittent, noisy, or absent due to transmission difficulties. Furthermore, at the time of JFG development, the proposed reference-signal-transmission path contained phase-locked oscillators that could introduce frequency offsets as great as several parts in 10^6 if they lost phase-lock. These phase-locked oscillators were not used in the final BSRFs distribution network. Because of these reference transmission difficulties, the JFG was designed to rely on only one characteristic of the reference signal—its average frequency as measured during times of good transmission.

The low drift rate of the 39A DCXO allows a JFG to ignore the reference during times of poor transmission. Although this drift rate is low, it is not zero and cannot be ignored. In the JFG the DCXO frequency-control word is held constant until the average of a large number (256 minimum) of short (16 seconds), valid frequency comparisons

between input and output signals indicates a frequency offset greater than $1/10^{10}$. Then the control word is changed to move the DCXO frequency $2/10^{10}$ in a direction opposite to the offset. This frequency-control operation can be separated into three functions: frequency-offset detection, averaging, and frequency correction.

3.2.1 Frequency-offset detection

Figure 5 shows, in general terms, the basic portion of the circuit by which the JFG detects small values of $\Delta F/F$ between two frequencies A and P. In the first portion of this circuit, divider D and the two mixer-bandpass filter circuits develop a frequency $A/D + \Delta F$. The final portion of the circuit consists of a frequency divider N whose output controls a gate, allowing the frequency A to be counted for one gate period, resulting in the count C. C_o is the count when $\Delta F = 0$. The count ΔC contains the $\Delta F/F$ information obtained in the time T.

Implementation of this circuit involves the selection of values for D and N and the number of stages in the counter. To maximize the $\Delta C/T$ ratio, D is made as large as possible within the constraints imposed by the bandpass filter requirements and the short-term frequency stability of the signals. In the JFG, $D = 320$; hence, $A/D = 64$ kHz. N is selected to provide the necessary precision (ΔC) in a reasonable time (T). N is also considered in the counter design. A counter large enough to count to C would be larger than necessary. In the JFG frequency-control circuit the largest $\Delta F/F$ that must be measured by the counter is $1/10^8$; this limits ΔC. If one selects N and the number of
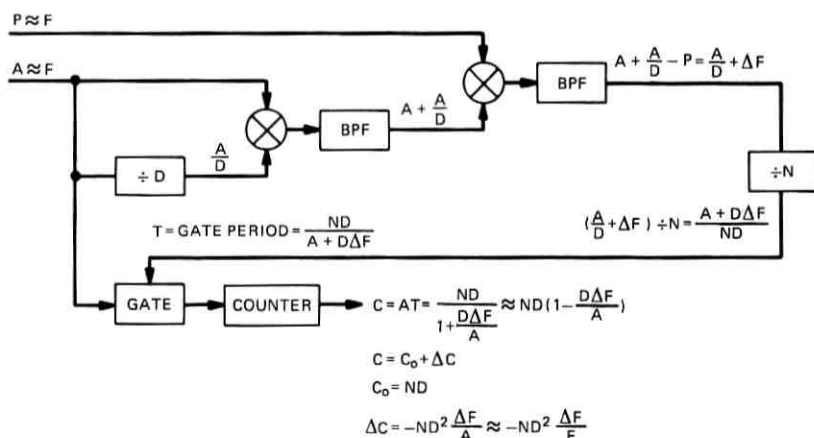

Fig. 5—Frequency-offset detection.

counter stages so that after multiple overflowings C_o leaves a remaining count of all zeros, the number of counter stages need only be sufficient to detect the sense and magnitude of ΔC. In the JFG, $N = 2^{20}$ and the counter has 19 stages. This results in a circuit capable of detecting sense and magnitude of frequency offsets in the $1/10^{11}$ to $2/10^6$ range.

Figure 6 shows the application of this circuit in the JFG. During the 16.384-second gate period provided by GATE 1, the COUNT signal (the JFG output frequency) is counted. At the end of that period, ΔC is examined and if it indicates a $\Delta F/F > 1/10^{10}$, high or low, a pulse is sent to the high or low input of the ACCUMULATOR. The INHIBIT 2 signal will interrupt this process and reset the circuit should other, quicker-acting circuits determine that the frequency comparison is not acceptable for frequency control.

3.2.2 Measurement averaging

Averaging of the results of the 16.384-second frequency comparisons is done in the ACCUMULATOR. This circuit is a nine-stage up-down binary counter. Starting from its midrange, or reset, position, this counter counts up or down depending on the sense of $\Delta F/F$. When this counter has received 256 more pulses in one direction than in the other, it sends two pulses to the MEMORY by way of the FREQ CONT MODE SELECTOR.

3.2.3 Frequency correction

The normal position of the FREQ CONT MODE SELECTOR will connect the ACCUMULATOR to the MEMORY. The MEMORY provides the 14-bit frequency-control word to the 39A DCXO. In the 39A a D/A converter plus varactor diode circuit provide a linear relationship between oscillator frequency and digital control word. The least significant bit (LSB) of the control word corresponds to $\Delta F/F = 5/10^{11}$; so the 14-bit word provides a frequency-control range of $\pm 4/10^7$. The MEMORY keeps the LSB at 0 and increases the higher bits by one for each pulse it receives. Thus, the two pulses from the ACCUMULATOR result in a $2/10^{10}$ frequency change.

The maximum correction rate of the frequency-control circuit is $2/10^{10}$ per 256 frequency comparisons made at a rate of one every 20.48 seconds; this converts to $3.4/10^9$ per day.

3.3 Frequency alarms

Figure 6 shows a number of frequency alarms being generated. The circuits generating these alarms are interconnected to utilize the inverse relation between the value of $\Delta F/F$ and the time needed to detect it.
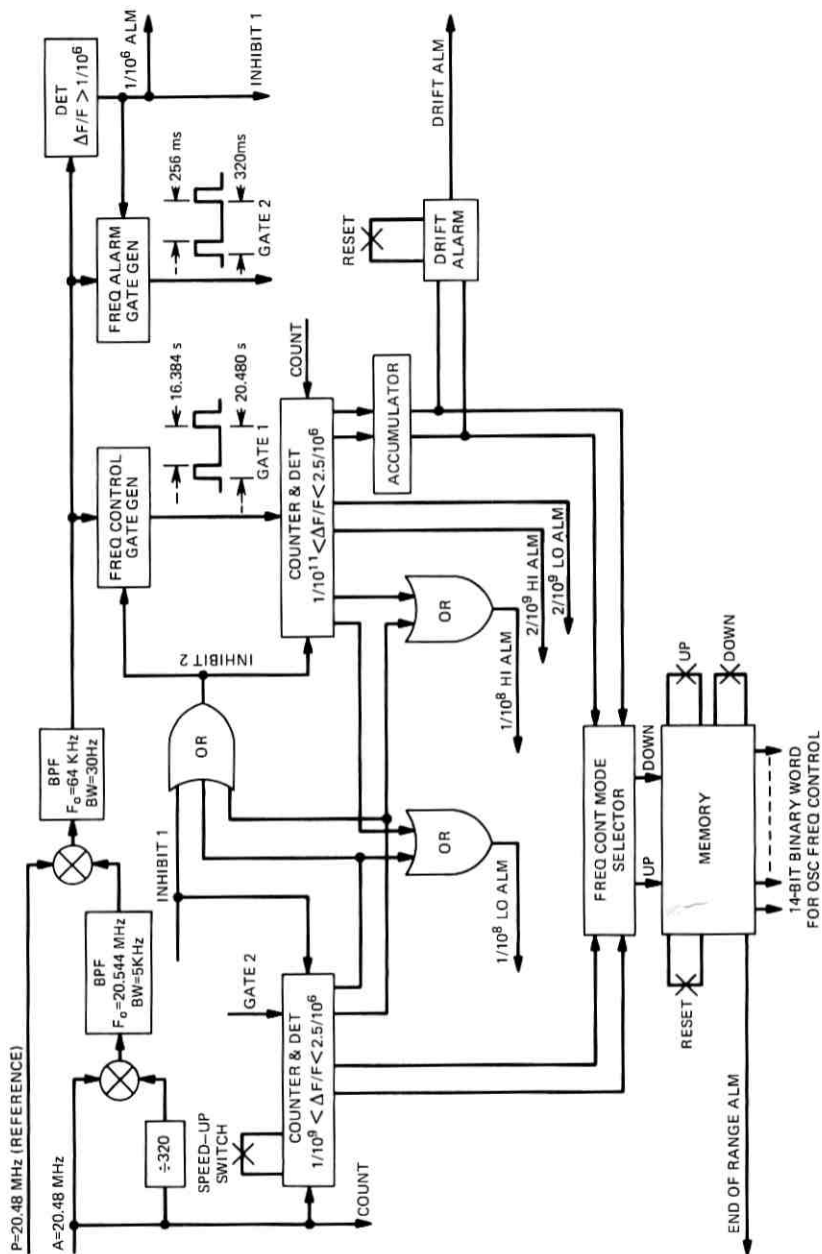
Fig. 6—Frequency-control-and-alarm circuit.

3.3.1 1/10⁶ alarm

This is the coarsest frequency alarm and utilizes the 30-Hz bandwidth of the 64-kHz bandpass filter. At the output of this filter a ΔF of 15 Hz or greater causes a 3-dB or greater drop in signal level. This drop corresponds to a $\Delta F/F$ of $7.5/10^7$ or greater. A level-detector circuit, DET, detects this and produces the $1/10^6$ ALM and INHIBIT 1 signals. (The $1/10^6$ designation is nominal.)

3.3.2 1/10⁸ alarms

GATE 2 is called the frequency-alarm gate, and the circuits generating it and using it comprise the frequency-alarm circuit. This circuit uses the frequency offset detection scheme shown in Fig. 5. $N = 2^{14}$ and the counter has 12 stages; so the circuit has a range from $1/10^9$ to $2.5/10^6$. The INHIBIT 1 signal interrupts and resets this circuit which makes a $\Delta F/F$ measurement every 320 ms. The main function of this circuit is to produce the $1/10^8$ HI ALM and $1/10^8$ LO ALM and shut down the frequency-control circuit when $\Delta F/F > 1/10^8$.

Since the counter used in the frequency-control circuit can detect $1/10^8$ offsets, it is also used to produce the $1/10^8$ HI ALM and $1/10^8$ LO ALM. This provides protection in case of frequency-alarm-circuit failure.

3.3.3 2/10⁹ alarms

The $2/10^9$ HI and LO alarms are produced in the frequency-control circuit. These alarms alert maintenance personnel to problems before they affect service.

3.3.4 Drift alarm

A frequency drift less than $3.4/10^9$ per day can be compensated for in the JFG. An oscillator with a drift rate greater than $3.4/10^9$ per day will eventually produce a frequency alarm. An oscillator with a lower drift rate will not cause such an alarm and no action would be taken until the oscillator reached the end of its control range and had to be replaced. The DRIFT ALARM forestalls this. The DRIFT ALARM circuit notes each frequency-correction pulse from the ACCUMULATOR, and generates the DRIFT ALM signal when the MEMORY has received 64 more pulses in one direction than in the other. The DRIFT ALM signal notifies maintenance personnel, who manually reset the circuit and record the alarm. The record of DRIFT ALM times and oscillator frequency-control words indicates oscillator drift rate.

3.3.5 End of range alarm

If the oscillator is at the end of its control range; i.e., its control word is all ones or all zeros, the MEMORY produces the END OF RANGE ALM and inhibits itself from further changes. The oscillator must then be replaced. However, long before this happens, maintenance personnel should have predicted it from the drift-alarm records.

3.4 Operational features of the JFG

Most functions within the JFG are automatic, but installation and maintenance operations must be performed manually. The JFG design has tried to facilitate these operations.

3.4.1 Rapid correction of frequency

At time of installation, or following a loss of power, the oscillator frequency might need to be changed so much that the $3.4/10^9$ per day rate would be far too slow. Shown in Figs. 6 and 7 are three features to provide more rapid frequency correction. First, there is the MEM RST switch which can put the frequency control word in its midrange position. Second, there are the $1/10^{10}$ UP and DOWN switches which change the oscillator frequency by $1/10^{10}$ with each operation. Third, there is the speed-up circuit which uses the 320-ms period of the
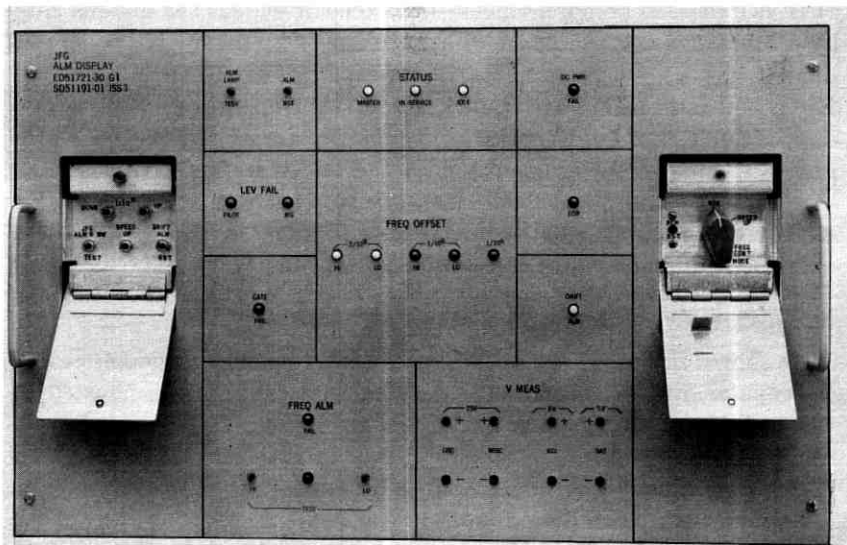

Fig. 7—Alarm-display panel for jumbogroup frequency generator.

frequency-alarm circuit to correct the oscillator at a $6/10^{10}$ per second rate. To operate the speed-up circuit, the FREQ CONT MODE switch is moved from the NOR to SPEED-UP position and then the SPEED-UP pushbutton switch is operated. The frequency-alarm circuit automatically stops sending pulses to the MEMORY when the $\Delta F/F$ is reduced below $1/10^9$. Figure 7 shows that all these controls are located behind covers to forestall their misuse. Note the angle bracket on the cover of the FREQ CONT MODE switch; this bracket forces one to return the switch to its NOR position before the cover can be closed.

3.4.2 Alarm lock-up

Intermittent alarms have always been the bane of maintenance personnel. In the JFG, alarm indications are held until they can be noted or until their cause has been corrected, whichever comes later. Figure 7 shows the ALM RST switch which will extinguish an alarm lamp if the trouble has gone. If the trouble still exists, the lamp cannot be extinguished. An exception to this lock-up is that all alarms associated with the JFS reference signal do not lock up.

3.4.3 Frequency alarm test

A failure of the frequency-alarm circuit could go unnoticed. The JFG provides a simple way of checking this. Operation of the FREQ ALM TEST switches, HI or LO, artificially shortens or widens the 256-ms gate by 1.6 ms, causing the effect of a $1/10^8$ HI or LO offset. If the circuit detects this, a green lamp lights; if it does not, the red FAIL lamp lights. No other indications are generated in this operation. But if one desires to check the interaction of this alarm circuit and the JFS alarm and switching circuits, operation of the JFS ALM & SW TEST switch (see Fig. 7) simultaneously with the HI or LO switch will allow $1/10^8$ alarms to be transmitted from the JFG to the JFS. Note that these operations do not disturb the JFG output frequency.

IV. FREQUENCY DIVIDER AND PHASE COMPARATOR

As shown in Fig. 1, the JFS contains two frequency dividers. Called the regular divider and standby divider, these units are connected to the outputs of the regular JFG and standby JFG, respectively. Each frequency divider provides as an output 1.024-MHz, 2.56-MHz, and 20.48-MHz signals multiplexed onto one coaxial cable.

Figure 8 is a block diagram of the basic circuit showing that phase control is provided for all frequencies. The 20.48-MHz phase-adjust networks provide a 180-degree control range. The fine phase adjust-

INPUT
20.48 MHz

DIVIDERS
WITH
FINE COARSE
PHASE PHASE
ADJ ADJ

LEVEL
ADJ

OUTPUT
20.48 MHz
2.56 MHz
1.024 MHz

Σ

÷8   BPF
      2.56 MHz

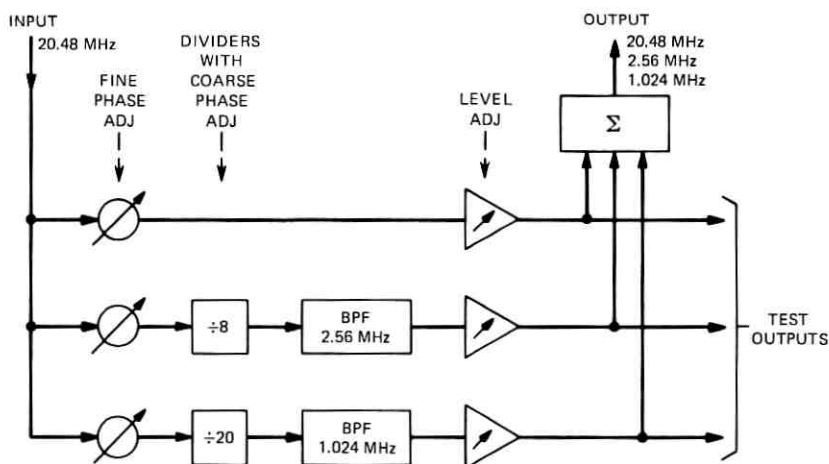÷20   BPF
      1.024 MHz

TEST
OUTPUTS

Fig. 8—Frequency divider.

ment for the 1.024-MHz and 2.56-MHz output signals is done at 20.48 MHz to avoid component problems associated with lower-frequency-phase-adjust networks. The 180-degree phase adjust at 20.48 MHz yields 22.5 degrees at 2.56 MHz and 9 degrees at 1.024 MHz. Coarse-phase-adjust circuits are included in the two dividers to increase the range of phase adjustment at the two lower frequencies. Each divider skips a single input pulse for each push of its associated pushbutton, thus producing a step change in its output phase. From each divider, the test outputs, shown in Fig. 8, are connected to the phase-comparator panel, which provides one meter for each frequency. A meter null is an indication of phase coincidence. All three meters must be nulled to indicate that the signals from the two dividers are in phase. In adjusting phase to produce these nulls, the coarse-phase adjust of the in-service divider should not be used.

V. PROTECTION SWITCHING

5.1 General description

Protection switching plays a vital role in ensuring the continuity of JFS output signals. Figure 9 shows the switching arrangement; the switches are fabricated with coaxial, dry-reed contacts. The normally closed contacts of these switches are magnetically biased so there is no need for switch coil current while the switch is released. Since the switch is normally released, a loss of switch power will normally not affect the JFS output signal. As shown in Fig. 9, with the switches
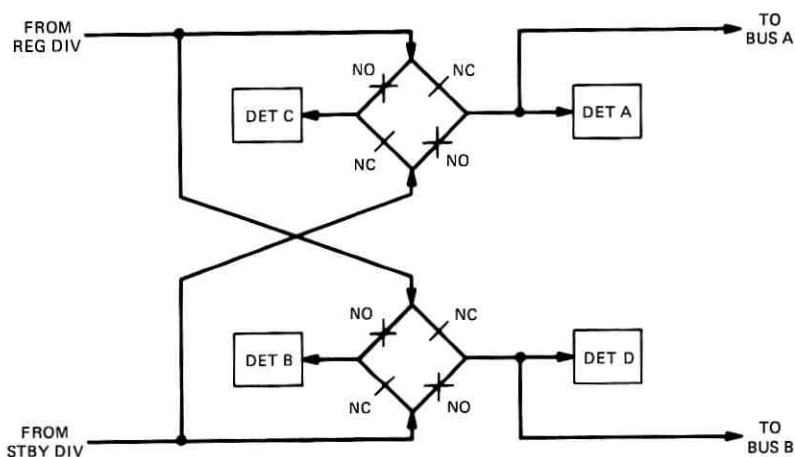
Fig. 9—Protection switch and level detectors.

released, the regular divider feeds detectors A and D and output buses A and B, while the standby divider feeds detectors B and C. When the switches are operated, the standby divider feeds detectors A and D and output buses A and B, while the regular divider feeds detectors B and C.

The switch-control circuit energizes (to operate) or deenergizes (to release) the switch coils in a "make before break" sequence. This feature has little importance if the switching is being done automatically due to a signal fault. However, if the switching is being done manually for any nonemergency reason, the switch input signals can be adjusted to be of equal phase and level; this combined with the "make before break" feature minimizes perturbations of the JFS output signal caused by nonemergency switch operation.

5.2 Manual control

Manual control is provided in the JFS alarm-and-switch-control panel by means of pushbuttons. Accidental manual switching is prevented by requiring multiple simultaneous pushbutton operations for switching. It is possible to override automatic-switch-control signals manually.

5.3 Automatic control

The alarms received from the four level detectors (Fig. 9) and the JFG's are processed by the switch-control circuit, which sends a com-

mand to the switches if an automatic switch is warranted. The switch will operate if detector A or D indicates an unusual change in signal level or if the regular JFG indicates it is impaired sufficiently to require being switched out of service. Should detector B or C indicate an unusual change in signal level or the standby JFG indicate it is impaired, switch operation will be inhibited.

5.4 Manual rearrangement

If a JFG failure occurs, the JFS may have to operate for some time with only two JFG's. Patching facilities are provided for restructuring the JFG interconnections (Fig. 10). For example, if the master JFG fails, the standby JFG temporarily serves the master JFG function. The alarm-and-switch-control circuit must be modified to recognize the new arrangement. This modification is achieved by selecting a position on a rotary switch corresponding to the new arrangement, and then operating an associated pushbutton switch. Because the JFS can function for some time with only two JFG's, the JFG is spared on a regional rather than a per-office basis. For most of the other units in the JFS, spares are included either in the JFS itself or in each office.

A jack-and-patch-plug arrangement has been provided in the alarm-and-switch-control panel so that the JFS output signals can be routed around the coaxial switches. Using this arrangement to take the switches out of service produces small phase and amplitude "hits" on
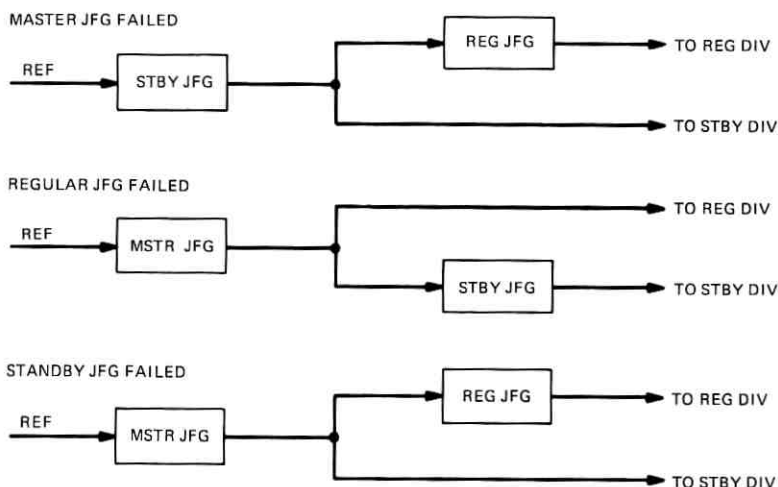

Fig. 10—Jumbogroup frequency supply operation using two jumbogroup frequency generators.

the JFS output signals. When the patching operation has been completed, the signal through the patched path is within 2 degrees and 0.2 dB of what it had been through the switch path.

A rotary-switch-and-power-meter arrangement has been provided on the alarm-and-switch-control panel to enable the craftsperson to read the level of each frequency (1.024, 2.56, and 20.48 MHz) at detectors A and B shown in Fig. 9.

VI. TEST UNIT

The JFS incorporates a test unit to measure small frequency offsets, since a typical office does not have precision-frequency-measuring equipment.

The test unit can compare two frequencies of approximately 20.48 MHz, and display on numeric indicators the fractional frequency offset in parts in 10^{10} or 10^{11}. The test unit has two modes of operation, either of which can be selected by a switch on the front of the panel. The 6 SEC, or normal, mode measures frequency differences of parts in 10^{10}. The 30 SEC, or extended precision, mode measures frequency differences of parts in 10^{11}. The frequency-offset information is displayed on LED numeric readouts.

Signals to be compared must have a level greater than −30 dBm into 75 ohms, and the frequency difference must be less than one part in 10^6. If either of these conditions is not met, the OUT OF RANGE indicator will light, and the test unit will not operate. The COUNTING indicator is lit while the test unit is counting. When the COUNTING indicator goes off, information is transferred from the counter to the numeric readouts. After approximately six seconds, a new count is taken.

The internal functions of the test unit can be divided into three sections: gate generator, counter, and display. The gate generator provides a gate whose period is a function of ΔF, the frequency difference between the two input signals. In the counter, the duration of the gate is measured by counting proportional to the fractional frequency offset $(\Delta F/F)$. This is similar to the frequency-offset-detection technique used in the JFG. The information in the detection counters is transferred to a digital display unit.

VII. CONCLUSION

The first JFS production unit was shipped from the factory in June 1973. There are presently more than ten JFS bays installed at L5 main

stations. The operational performance of these units has been excellent, and they have met all of their design objectives.

The availability of these accurate and reliable frequency supplies has stimulated the modernization of the entire Bell System synchronization network.[4] Beginning in 1974, the United States will be divided into approximately twenty synchronization regions, and telephone offices in each region will receive their synchronization signals from a regional frequency supply (RFS). Certain L5 main stations will become regional sync centers, with the L5 JFS serving as the RFS and providing 64-kHz and 512-kHz sync signals to telephone offices in its region. In areas where no L5 facility exists, a special two-JFG version of the JFS will serve as the RFS. Thus, the entire frequency-division multiplex plant, including both radio and coaxial transmission systems, will soon be synchronized through JFS equipment.

In addition to serving L5 and the Bell System synchronization network, the JFS has sufficient frequency precision to serve future systems having ten times the bandwidth of L5.

VIII. ACKNOWLEDGMENTS

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L5 SYSTEM:

Signal Administration and Interconnection

By R. K. BATES and D. J. ZORN

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Each L5 coaxial line is capable of carrying three 3600-channel basic jumbogroup signals that are translated to and from the L5 line spectrum (3.124 to 60.556 MHz) through jumbogroup multiplex equipment. The L5 line signal also includes several line pilots and switching, maintenance, and reference signals. Administration of all these signals is performed by the L5 line connecting circuits, which vary in complexity with the type of main station they serve.

In addition to the circuits required to handle the various components of the L5 line spectrum, signal administration is also required at basic jumbogroup frequencies, before the jumbogroup multiplexing step, to allow interconnection to lower-order multiplex or other systems using the basic jumbogroup frequency format of 0.564 to 17.548 MHz. To perform this function, the basic jumbogroup trunk bay was developed and provides interconnection flexibility hitherto unavailable for direct connections to other long-haul systems, such as L4, LMD, TD, or TH radio, or L5 systems of another route.

I. INTRODUCTION

The L5 coaxial line is a transmission facility with a message capacity of three jumbogroups. These jumbogroups are placed on the line in the frequency format shown in Fig. 1. Each jumbogroup begins as a basic jumbogroup signal, formed by the mastergroup multiplex-2 (MMX-2) frequency-division multiplex or the basic jumbogroup trunk equipment. This basic jumbogroup consists of six 600-channel basic mastergroups and has a frequency assignment identical to that of the L4 line assignment.[1] Each of three 3600-channel basic jumbogroups is translated to the L5 line spectrum through the jumbogroup multiplex
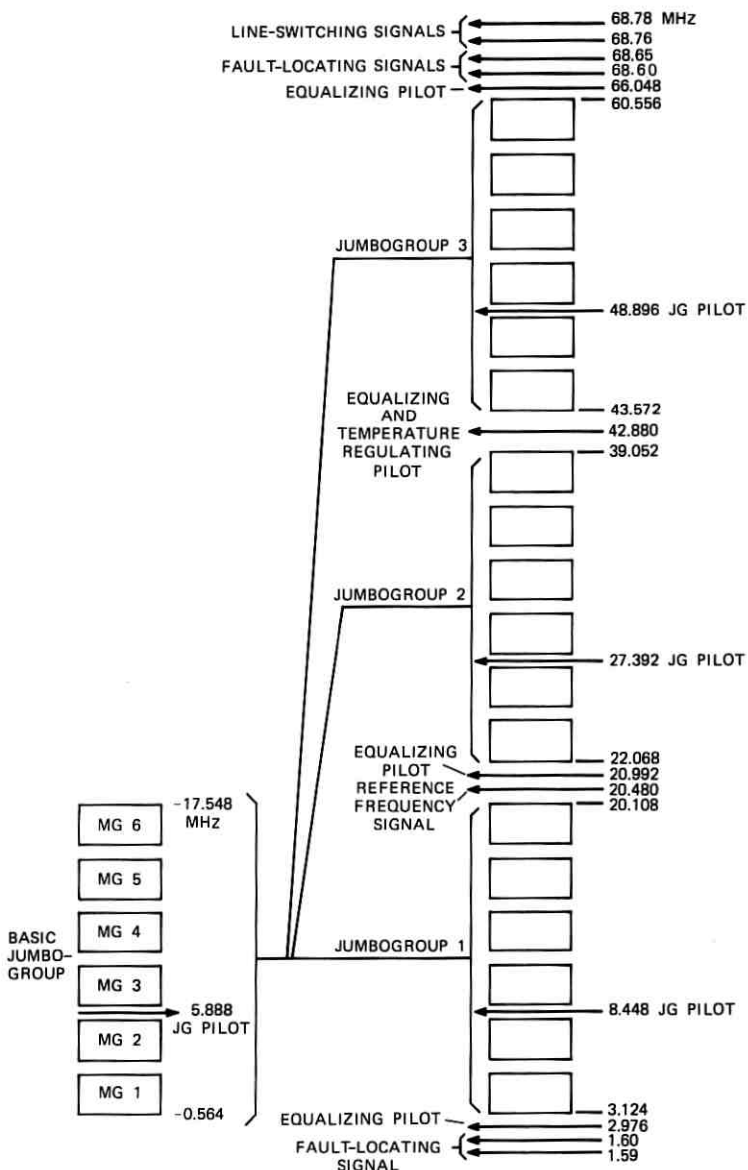
Fig. 1—Three-jumbogroup line frequency spectrum.

(JMX) frequency-division multiplex equipment[2] to form the 10,800-channel line signal.

In addition to the message band, Fig. 1 also shows the placement of

several pilots used for the dynamic equalization and temperature regulation of the line,[3] switching signals required for the control of the line-protection switching system-3 (LPSS-3),[4] fault-locating signals used for line maintenance tests in the transmission surveillance system (TSS),[5] and a reference frequency signal used by the jumbogroup frequency supply (JFS).[6] Several rules and administrative functions must be applied to the message and other signals forming the L5 line spectrum which vary under specific situations. The line-connecting circuits described in Section II perform all of these functions.

In the evolution of the various long-haul systems such as the L-carrier and TD and TH radio systems, the need for more direct interconnection between these systems in large channel blocks became increasingly evident. Prior to the development of L5, the predominance of intersystem message interconnection was done on a basic mastergroup basis,* which requires the use of costly MMX terminals and mastergroup connectors. The need for a simpler and less expensive means of intersystem interconnection was recognized early in the L5 development, especially in view of the large 180-mastergroup capacity of this new system.

This need was met with a new bay, designated the basic jumbogroup trunk bay (BJGT), which allows interconnection of single- or multi-mastergroup signals in the basic jumbogroup spectrum; i.e., before the jumbogroup multiplexing step to the L5 spectrum through JMX equipment. Interconnections may be made directly to radio systems (using 3A wire line entrance links), L4 systems, L-carrier mastergroup digital (LMD) terminals, MMX-2 terminals, or other L5 systems. The BJGT circuits are covered in detail in Section III.

II. LINE-CONNECTING CIRCUITS

As mentioned in the introduction, the function of the line-connecting circuit is to process the L5 line signal in accordance with circuit requirements and certain administrative rules. Although many options are required to handle the various conditions that arise, there are only three basic line-connecting arrangements, one for power-feed main (PFM) stations, another for switching power-feed main (SPFM) stations and, finally, one to cover terminal stations or terminal main (TM) stations.†

All line-connecting equipment is located in the line transmit-receive bay. With the exception of line-connecting equipment, transmit-

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* The basic mastergroup is the "U600" output of the L-type multiplex terminal (Ref. 7).
† The distinction among the four types of main stations is covered in Ref. 8.
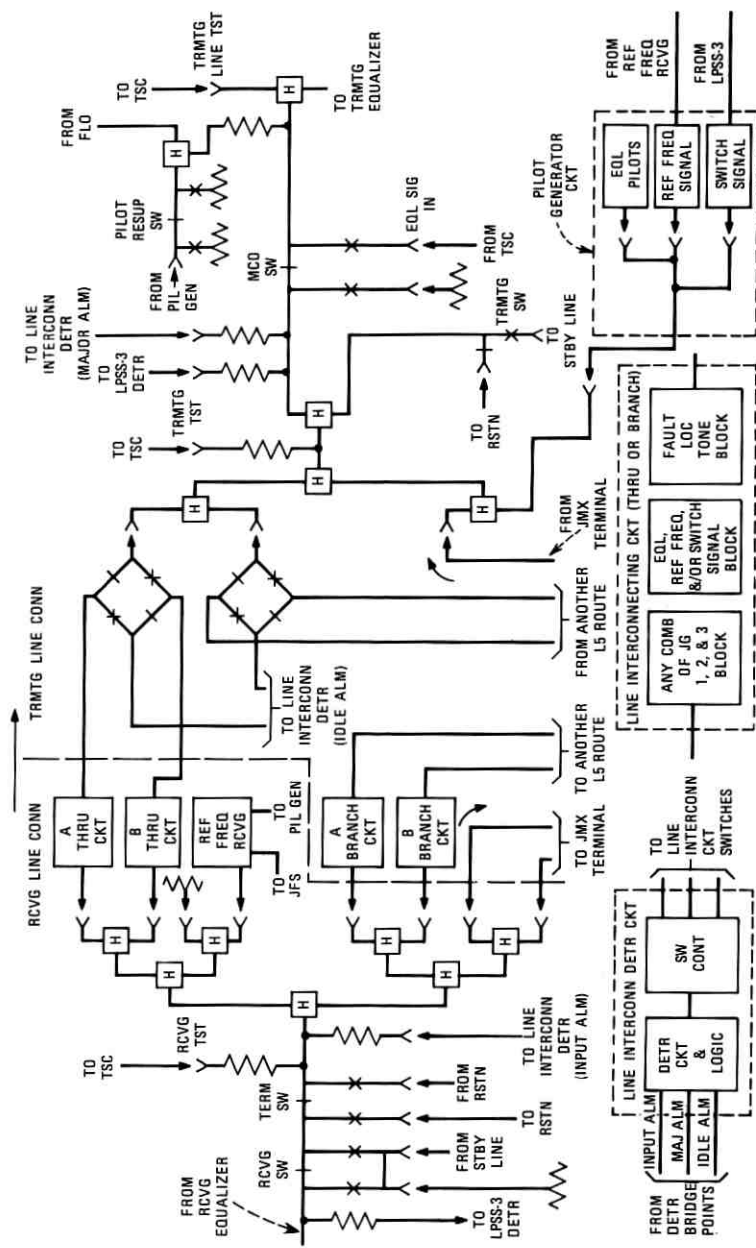
Fig. 2—Simplified block diagram of TM station L5 line-connecting circuits.
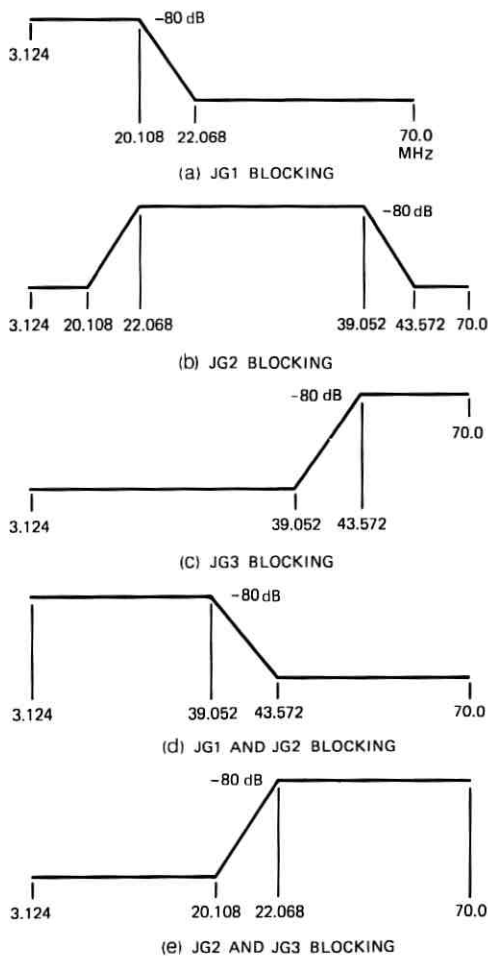
Fig. 3—Jumbogroup blocking filter requirements (frequency in MHz).

receive bays are virtually identical for all types of stations.[3,9] The line-connecting circuits are simplest at PFM stations where no signal processing is performed, and all message continues through the office. A pad and hybrid, for fault-location oscillator (FLO) access to the line, are the only apparatus in the line-connecting circuit. At SPFM stations, all message is again connected through; however, switching apparatus, filtering, and additional access to the transmission surveillance center (TSC) are added to the line-connecting circuits for system administration.

Terminal or TM stations require similar basic equipment units in the line-connecting circuits, but the terminal station options within that equipment are simpler. Each TM station transmit-receive bay requires six basic circuits to perform the line-connecting functions: the receiving line-connecting circuit, transmitting line-connecting circuit, pilot generator circuit, line-interconnecting circuits, line-interconnecting detector circuit, and, optionally, the line-branching circuit. These circuits are shown in the block diagram found in Fig. 2, and are discussed more thoroughly below.

The signal administration role of the line-connecting circuits in TM stations may be reduced to six major functions:

(*i*) *Message Administration:* Provides routing of the message signal for any combination of the three jumbogroups.

(*ii*) *Pilot Administration:* Introduces equalizing and temperature pilots to the line and provides pilot blocking when required.

(*iii*) *Switching Administration:* Provides line switches, introduces switching signals to the line, and provides detection access and signal blocking when required.

(*iv*) *Transmission Surveillance Administration:* Provides access to the line at strategic points for TSC analysis, and furnishes signal blocking when required.

(*v*) *Reference Frequency Signal Administration:* Introduces a reference frequency signal to the line, and provides signal blocking when required and a distribution network for circuits requiring the use of a reference frequency.

(*vi*) *Restoration Access:* Provides access to the regular transmitting and receiving lines for restoring L5 over other facilities and access to the standby transmitting and receiving lines for restoring other facilities over L5.

2.1 Message administration

The simplest of the TM station line-connecting circuits is required when all message passes through the office; i.e., no jumbogroups are branched to other L5 lines or are dropped to JMX equipment. In such a case, the line signal entering the line-connecting circuit from the receiving equalizer (Fig. 2) would only connect (via splitting hybrids) to the A and B THRU CKT modules at the output of the receiving line-connecting circuit. No message blocking filters are required in the THRU CKT modules for this application.*

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* Although no message blocking filters are required, other circuitry may be required for other line-connecting functions at TM stations, as described below.

In the example shown in Fig. 2, the message administration is arranged to split the jumbogroup signals into three directions: one jumbogroup connected through, one branched, and one dropped to JMX. Assuming JG1 is connected through, JG2 branched, and JG3 dropped, blocking filters would be assigned to block JG2 and JG3 in the THRU CKT and JG1 and JG3 in the BRANCH CKT. No blocking filters are required in the drop circuit, since the JMX circuitry selects that jumbogroup for which it is equipped.

Any combination of jumbogroups can be passed in the through, branched, or dropped path. The blocking is accomplished through the use of five arrangements of filter designs developed for L5 use. The requirements for jumbogroup blocking are outlined in Fig. 3.* In all but jumbogroup 2 blocking, single high-pass or low-pass filter designs are used; however, in the case of jumbogroup 2, filters *d* and *e* are paralleled with a "split-apart" filter at the input and output of the paralleled pair. The advantage of split-apart filters, as compared to hybrids, allows better return loss with lower in-band loss.

A minimum of 80-dB out-of-band discrimination for each blocking arrangement is required. Return loss is approximately 26 dB (75 ohms). Maximum insertion loss is less than 4.0 dB.

2.2 Pilot administration

Four full-time pilots are required for the administration of the L5 line: 2.976, 20.992, 42.880, and 66.048 MHz. The 42.880-MHz pilot is used for dynamic equalization and temperature regulation as well as the switch initiation described below, whereas the remaining pilots are used for dynamic equalization only (see Fig. 1).

Ideally, it is desirable to maintain continuous-line pilot continuity throughout a frogging[8] section (approximately 800 miles). Two factors, however, preclude this possibility. When jumbogroup blocking filters are required at TM stations, one or more line pilots are attenuated to some degree because they fall within the attenuation region of the filters. In addition, the stability of each line pilot is adversely affected as the number increases of regulating repeaters or dynamic equalizers through which they must pass. Therefore, the pilots are blocked and reinserted at specific intervals.[3]

Since the administration of individual pilot blocking and reinsertion would be unwieldy, all pilots are blocked when any pilot or jumbogroup signal is blocked. Thus, rules for pilot blocking were established such

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* A more detailed description of filter design techniques for L5 is given in Ref. 10.

that all pilots are blocked when either (*i*) jumbogroup blocking is provided or (*ii*) pilots will otherwise pass through more than four E3 equalizers.

Minimum requirements for the pilot band-elimination filters are as follows:

Frequency ($\pm1/10^5$ from nominal)	Discrimination (dB)
2.976	50
20.992	50
42.880	80
66.048	50

Return loss requirements are 26 dB (75 ohms) and insertion loss is less than 4 dB.

2.3 Switching administration

The line-connecting circuits provide several functions for the administration of the LPSS-3. A bridged access point is provided at two locations, (*i*) on the line side of the receiving switch and (*ii*) on the line side of the transmitting-line switch hybrid (Fig. 2). The switch-initiator circuit monitors the 42.880-MHz pilot at these points for LPSS-3 operations.

Unlike L3 and L4, the L5 line switches are located within the line-connecting equipment to reduce the cable length in the through signal path. This was made possible through the use of parallel switch cabling, as opposed to the series arrangement in L3 and L4 systems.[4]

Two switches are provided in the receiving line-connecting circuits, designated the RCVG SW and the TERM SW. The RCVG SW furnishes access from the standby line when the associated regular line is out of service for any reason. This switch also provides a termination to the out-of-service line. The TERM SW is either automatically operated as a result of a line overload condition, thereby preventing noise propagation to other systems, or manually operated under certain abnormal line conditions.[4] The transmitting line-connecting circuit also includes two switches, the TRMTG SW and the MCO (message cutoff) SW. The TRMTG SW provides access to the standby line when the associated regular line is out of service. The MCO SW, also controlled by the LPSS-3, provides a means for opening the transmitting end of a regular line, primarily as an access for line equalization.

As indicated in Fig. 1, the line-switching signals occupy the band-width between 68.76 and 68.78 MHz. These signals are blocked in the

line-interconnecting circuit at every switching station to prevent interference with switching functions between switching sections.

In addition to those switches described for the regular bay, the standby transmit-receive bay line-connecting circuits use a receiving and transmitting line director switch (see Fig. 4).* These switches provide inputs and outputs to the 10 regular lines for switched access to the standby line.

2.4 Transmission surveillance system administration

Each transmit-receive bay at TM and SPFM stations has six measurement points which can be accessed by the TSC for automated remote measurements of the L5 line status.[5] In addition to the measurement points, an input access is also provided in the transmitting line-connecting circuit so that out-of-service frequency characteristic measurements may be remotely performed by the TSC on a switching section. This point, equalizing signal in (EQL SIG IN), is also used for line equalization described in an earlier section.

Another transmission surveillance system administrative function performed by the line-connecting circuits is that of fault-locating signal blocking. Fault-locating signals at 1.59, 1.60, 68.6, and 68.65 MHz must be blocked at each switching section to avoid interference with fault-location routines in other switching sections. A high-pass filter blocks the low-frequency fault-locating tones a minimum of 50 dB. An alternate high-pass filter design will also block the 2.976-MHz equalizing pilot (Section 2.2) when required.

Since switching (Section 2.3) and fault-locating signals are always blocked at each switching station, a single filter accomplishes both functions. Because of the relative closeness of the 66.048 equalizing pilot—which is not always blocked concurrently with the switching and fault-locating signals—a low-pass filter was not a practicable design. Instead, a band-elimination filter with a minimum of 50 dB of suppression was developed for this purpose. In those cases where the 66.048-MHz pilot is also blocked (Section 2.2), a low-pass filter is furnished to block all signals above the message band.

2.5 Reference frequency signal administration

The JMX requires a very stable frequency for the synchronization of its carrier supply circuits. This frequency, 20.480 MHz, is generated by the Bell System Reference Frequency Standard (BSRFS) located at

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* Figure 4 represents a simplified block diagram of the line-connecting circuits for a SPFM station. The circuitry is essentially similar to the TM station, with the exception of the line-interconnecting circuits, which are much less complicated.
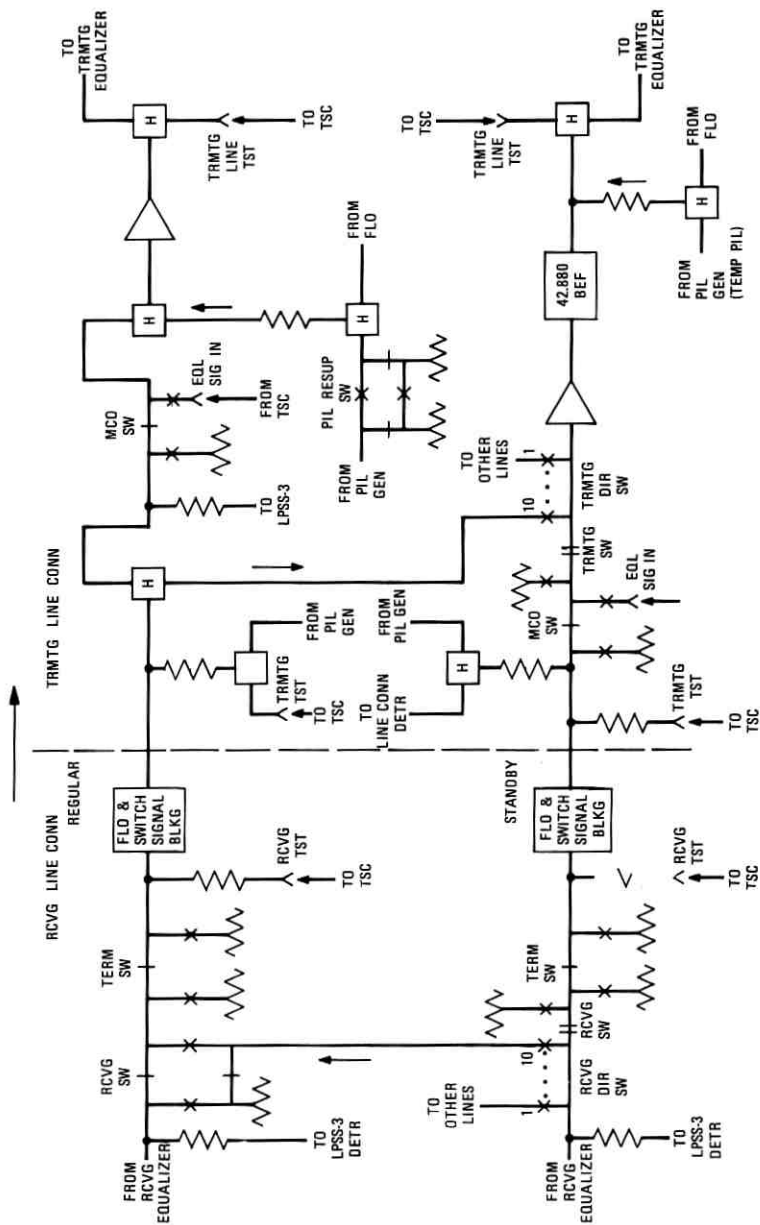
Fig. 4—Simplified block diagram of SPFM station line-connecting circuits.

Hillsboro, Mo.[6] The BSRFS will supply the reference frequency signal for all L5 systems, either directly, as in the case of the initial L5 Lillyville–Hillsboro route, or indirectly via other long-haul systems in subsequent routes.

Since the reference frequency signal is only required on the first regular working line and in only one direction, the line-connecting circuit administration of that signal is limited to that line. As in the case of the equalizing pilots, it is desirable to maintain the integrity of the reference signal throughout the route without restriction; however, if jumbogroup 1 or 2 is blocked in the through route, some attenuation of the signal occurs that requires that a 20.480-MHz band elimination filter be added to the receiving circuit of the through route to block the signal sufficiently to prevent its interfering with the reference signal reinserted in the transmitting line-connecting circuits of that route. In such cases, the 20.480 signal is bypassed around the through-line-connecting circuit, as described in the next paragraph. The characteristics of the filter are similar to those of the 20.992-MHz band-elimination filter described in Section 2.2.

In those offices requiring a reference signal for JMX terminals (via the JFS circuitry), for reinsertion in a through route or for insertion on an L5 branch route, a synchronizing receiving panel is provided. This panel, which accepts the entire line signal from a port in the receiving line-connecting splitting-hybrid circuit (Fig. 2), is furnished with a 20.480-MHz bandpass filter, amplifiers, an adjustable attenuator, and a hybrid tree that provides eight outputs for distribution as required. The bandpass filter provides a minimum of 80-dB discrimination approximately ± 75 kHz from the reference frequency signal.

2.6 Restoration access

The line-connecting circuits provide access for message service restoration both for L5 restoration and for the restoration of other long-haul systems over standby L5 lines.

Access to restore a failed L5 facility is provided at each regular line at the TERM SW (Fig. 2) in the receiving line-connecting circuit and the TRMTG SW in the transmitting line-connecting circuit. These access points connect—via the restoration patch bay—to the standby L5 line of another facility or, through idle JMX terminals, to spare equipment of other long-haul facilities. Another access point is provided at the TERM SW to allow monitoring of the failed lines at the restoration patch bay. Similar points are provided with the standby line connecting

equipment to allow direct restoration access to other failed L5 facilities or other types of long-haul systems through idle JMX equipment.

2.7 Line-interconnecting modules

During the early development of the L5 line-connecting circuits, it was recognized that considerable gain would be required at TM stations to offset the loss of the filters required for specific administrative functions. Since as many as 18 mastergroups may be served on each line-connecting circuit, redundancy was provided to assure reliability consistent with the overall L5 system. Although without redundancy the reliability of the amplifiers in isolation was considered adequate, the associated power, fusing, and related cabling associated with the active elements made redundancy necessary.

A beneficial byproduct of the redundancy in line-connecting circuits is the ability to provide all optional equipment in a modular package, thereby allowing in many cases in-service rearrangements of circuit assignments with less danger of service outages.

The line-interconnecting circuit modules are depicted in Fig. 2 as the A and B THRU CKT or A and B BRANCH CKT. The A and B paths terminate in a coaxial switch that provides the automatic protection for the A or B unit, whichever is serving the line. Jumbogroup pilots control the switch logic circuitry, which can be rather complex because of the wide number of options available in the line-connecting circuits. Three points of detection are required to satisfactorily implement the switching logic: (i) beyond the output of the switch to indicate a major alarm* at loss of pilot; (ii) at the output of the alternate leg of the switch to monitor the condition of the idle line-connecting circuit; and (iii) at the input of the modules to assure that the jumbogroup pilots are present at the proper level. Since as many as three sets of line-interconnecting modules may be required, i.e., one through circuit and two branch circuits, up to nine detector circuits are required as a maximum for monitoring the three jumbogroup pilots at the three points of detection. The simplest case, of course, is when all jumbogroups are "through," in which case only one jumbogroup pilot need be detected (at the three points) for the full message load.

Line-interconnecting circuits for the standby line are considerably simpler than the regular lines, since no jumbogroup blocking or synchronization administration is required. All intermediate stations are therefore "through routes" on the spare line for an entire frogging[8]

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* All line-connecting alarms and status indications may be connected to remote alarm systems.
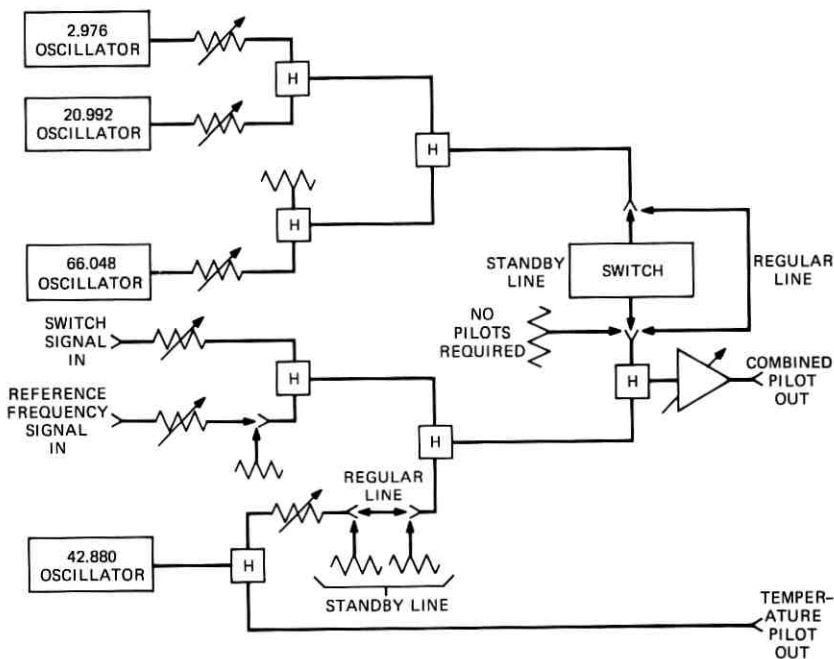
Fig. 5—Pilot generator circuit.

section. Although active circuitry is furnished with the standby line-connecting circuits, no automatic protection is required, inasmuch as service does not normally traverse an office on the standby line. An alarm is provided, however, to monitor continuity through the standby line-interconnecting module, and a spare module furnished may be manually patched in to replace it.

2.8 Pilot generator circuit

As shown in Fig. 2, the pilot generator circuit combines all the pilots and maintenance signals so that only one input port is required for their application to the transmitting line-connecting circuit. Several options within the generator, as noted in Fig. 5, are consistent with the various options required for line-connecting circuit administration.

The pilot-generator circuit provides the means for combining the switch signals from the LPSS-3 system, the reference signal from either the local standard or the receiving reference circuit (Section 2.5), and the four locally generated equalizing pilots. The four pilots are derived from free-running crystal oscillators mounted within the generator

panel, which are designed to maintain a frequency accuracy of approximately one part in 10^6 per year under normal office temperature variations. Harmonic distortion is required to be at least -40 dB relative to the fundamental.

In all SPFM and TM stations,* the minimum options required are the 42.880-MHz oscillator and the switch signal combining circuits. The 42.880-MHz signal is required for regular lines to connect to the pilot resupply switch (Fig. 2), which automatically reinserts the pilot should (*i*) the pilot be lost in the through path, or (*ii*) the MCO switch be operated for equalization or TSC measurements, thereby sustaining temperature regulation of the line during out-of-service conditions. Since the 42.880-MHz pilot for standby lines is always blocked and reinserted (Fig. 4), the oscillator is always required with standby transmit-receive bay pilot generator panels.

Another feature provided for standby lines is a switch for reinserting three of the equalizing pilots. This switch is provided at SPFM and TM stations where pilots are not blocked in the through path. Since there may be times when the standby line is being used for protection for sustained periods, the stations with through pilots would otherwise be without dynamic equalization; thus, when a receiving line switch occurs, three equalizing pilots are automatically reinserted for the duration of the switch.

III. BASIC JUMBOGROUP TRUNK CIRCUITS

As discussed in the introduction, the basic jumbogroup trunk bay design is unique in long-haul system development in that it provides a direct standard interconnection between the JMX and certain multiplex terminals and systems of multi-mastergroup capacity. Any combination of contiguous mastergroups in the basic jumbogroup format may be interconnected from L4 lines, MMX-2 or LMD terminals, radio systems using 3A wire line entrance links, or other JMX terminals as indicated in Fig. 6a.

Actually, a maximum of four inputs per jumbogroup is available to the BJGT bay, so that a more practical arrangement with contiguous mastergroup assignments is shown in Fig. 6b. Through judicious planning of the mastergroup assignments, line engineering personnel can more economically engineer facilities; however, there will still be cases where some interconnecting systems will carry identically

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* PFM stations are not furnished with pilot generator panels; however, a special portable, equalizing-pilot generator unit is available for installation and maintenance tests.

* UP TO FOUR MAY BE CONNECTED TO ONE JUMBOGROUP

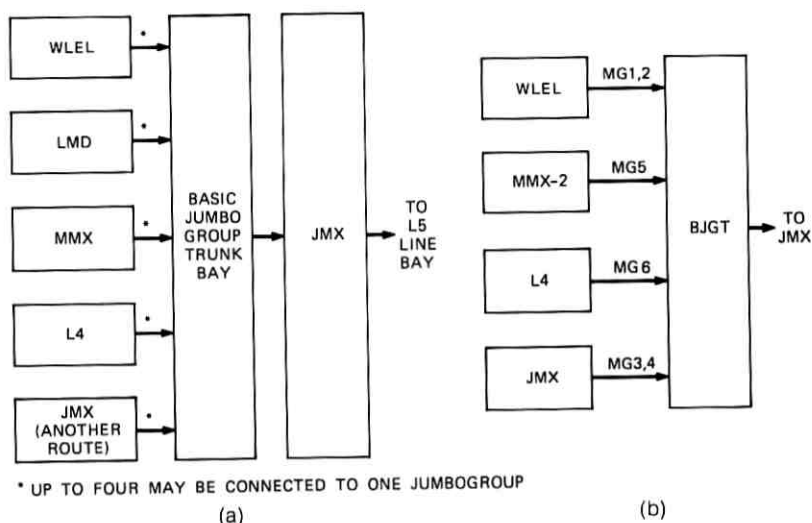(a)   (b)

FIG. 6—(a) Basic jumbogroup trunk function. (b) Example of basic jumbogroup trunk arrangement with four-system input.

numbered mastergroups, thus requiring multiplex equipment to reassign their frequency allocations.

The BJGT bay is required whenever a JMX terminal is furnished, since it provides not only the transmitting and receiving trunk circuits for interconnecting the JMX to the various terminal arrangements, but also the 5.888-MHz jumbogroup pilot. The BJGT bay provides more than 250 different list structures to accommodate all combinations of contiguous mastergroups to associated systems.

The transmitting and receiving trunk circuits consist of signal-processing panels and switch and logic panels. An example of the transmitting trunk circuit (transmitting to JMX), shown with a two-input circuit arrangement, is given in Fig. 7a. The signal from MMX, radio, LMD, L4, or another JMX terminal is split into two paths to identical, redundant signal processing circuits consisting of filters,* amplifiers, cable equalizers, and, in the case of L4, de-emphasis. Each processing circuit connects to a coaxial switch controlled by the logic circuitry in the switch and logic panel. The logic circuitry is activated when the mastergroup pilot level associated with the highest-numbered mastergroup in that particular circuit arrangement changes by a

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* The filters are low-pass, high-pass combinations, similar to those used in L4 blocking and branching arrangements,[1] and are required for removing unwanted signals in the various spectra of interconnecting systems.

(a)
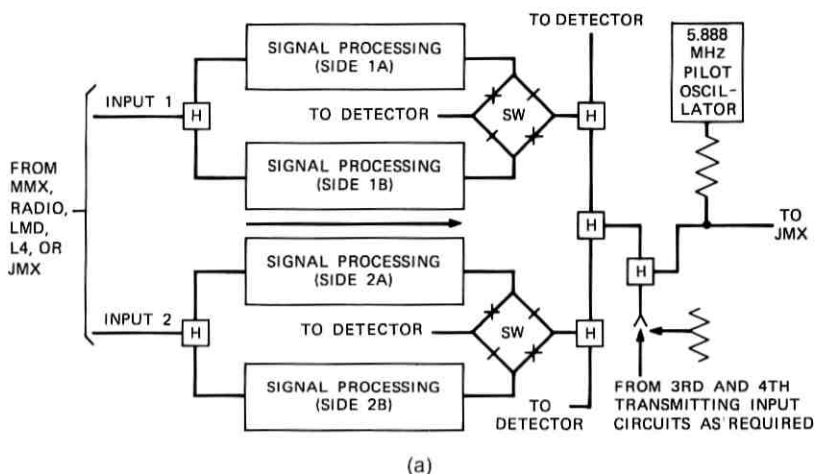
Fig. 7a—Transmitting basic jumbogroup trunk circuit.

(b)

Fig. 7b—Receiving basic jumbogroup trunk circuit.

predetermined amount. In addition, the logic circuitry provides local and remote alarming.

The transmitting trunk circuit also contains the 5.888-MHz basic jumbogroup pilot oscillator. The insertion of the jumbogroup pilot at the BJGT bay allows the interbay trunk to the JMX bay to be alarmed, thereby offering further protection against personnel error or cabling problems.

The receiving basic jumbogroup trunk circuits are very similar in design to the transmitting circuits, as shown in Fig. 7b; however, in addition, a 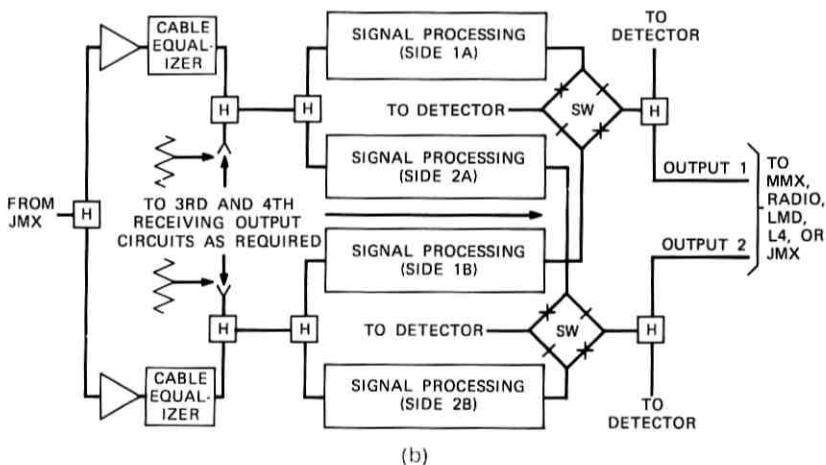5.888-MHz band elimination filter is provided to prevent interference with connecting systems, and pre-emphasis is furnished when connecting to L4.

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L5 SYSTEM:

Physical Design

By R. J. WIRTZ and W. G. ALBERT

The L5 Coaxial-Carrier Transmission System equipment was designed to provide long-term reliable service in diverse environments. The line repeaters, for instance, are housed in apparatus cases located in manholes that are subjected to flooding, whereas the main-station repeaters and terminal multiplexing equipment are located in the controlled environment of underground or aboveground buildings, often only partially attended by maintenance personnel. In addition to describing novel physical designs, this paper covers other very important considerations such as thermal design, manufacturability, incorporation of hybrid integrated-circuit technology, efficient "building-block" system growth capability, and long-term reliability.

I. INTRODUCTION

1.1 Outside plant design considerations

Proposed approaches to the physical design of L5 repeaters covered a wide array of possibilities ranging from the buried repeater concept to that of an ideal conversion. The former would capitalize on the high reliability and trouble-free performance of solid-state repeaters to gain cost advantages in outside plant for new routes, but would require initially equipping all tubes of a coaxial cable with repeaters. The latter approach would feature repeaters configured to closely resemble existing L4 system repeaters.[1] It would encourage conversions from the L4 to the L5 system and take advantage of as much existing outside plant as possible. The range of proposals also included such alternatives as limited-access enclosures, underground lockers, and single-repeater apparatus cases, all of which would have required extensive outside-plant apparatus development.

Experience with the L4 system showed that accessibility to line equipment is essential during initial installation of a system. This precluded burying L5. Perhaps, after a shakedown period, buried repeaters can be developed for new-route applications. One-mile repeater spacing and continuing improvements in reliability of solid-state devices may create a wish to bury L5 in the future. In this case, the basic repeater, the least complicated and most numerous of the L5 repeaters, is the most likely first candidate.

Clearly, the pursuit of a conversion design was the most logical initial approach to the physical design of the L5 repeaters. There will certainly be a continuing need to convert L4 routes to L5 when the demand for circuits exceeds L4 route capacity. Also, the L5 trial in mid-1970 was only realized by the use of existing plant designs where possible. In addition, by pursuing a conversion design, the development of the L5 system did not depend first on the acquisition of knowledge of a wholly new repeater environment. The existing L4 environment was usable for evaluation of the L5 repeaters during the development period. Conversely, it was also understood that pursuing a conversion design would not preclude the use of newly developed L5 outside plant in applications at the 1-mile repeater points where new manholes are required, and along new routes, if the need for new plant could be justified. Thus, the established schedules for development, trial, and commercial service for the L5 system, together with the economic advantages of reusing existing outside plant, all favored the conversion-design approach to L5 physical design.

An early plan was directed toward ideal conversion—the situation in which the L5 repeater shape would correspond closely to that of the L4 repeater, allowing easy replacement in the existing L4 apparatus case housing and chassis assembly. However, the proposed L5 system fault-location scheme requires 14 copper-wire pairs in the coaxial cable, whereas the L4 fault-location scheme requires only 5 wire pairs. This became the controlling factor in causing a change in the L4 apparatus case chassis. Once it was recognized that a chassis change was needed, the degree of change was academic, and a modified conversion plan was adopted. This plan was directed toward developing a repeater configuration that would still fit the L4 apparatus case, but required that the chassis of the apparatus case be changed to accommodate administration of the L5 fault-location pairs, to promote good heat transfer, and to simplify repeater installation. Since the L5 basic repeater dissipates some 18 watts (compared to 13.5 watts for its L4
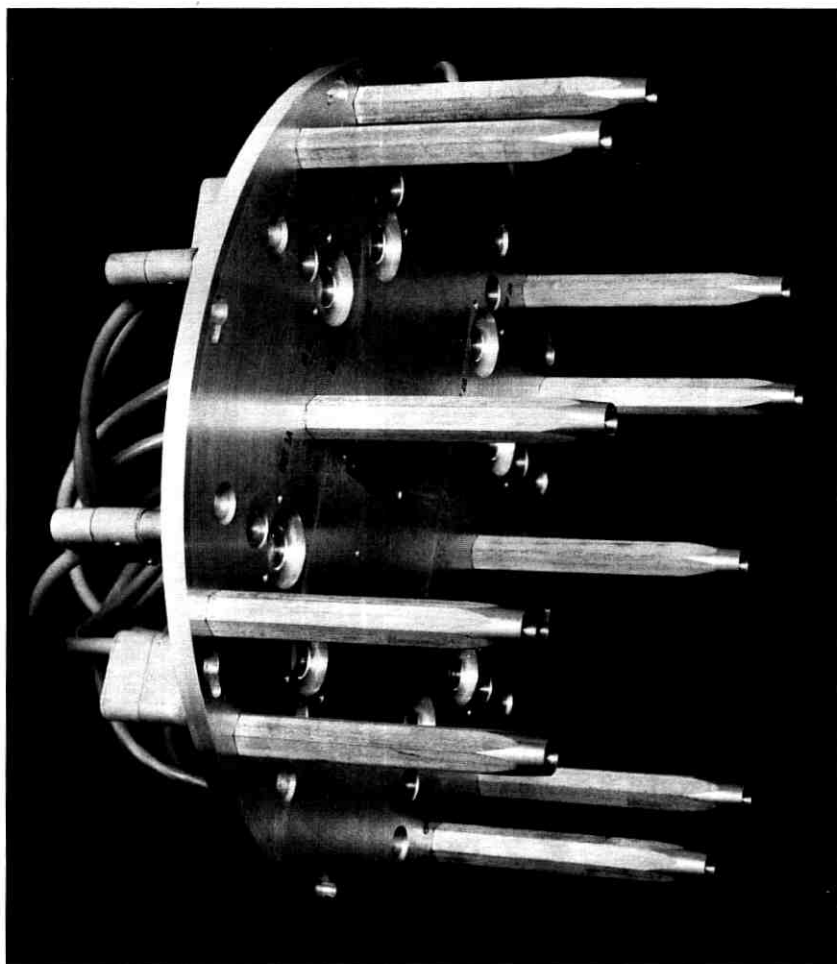
Fig. 1—Apparatus case chassis.

counterpart), it was necessary to design a more efficient thermal circuit in the L5 repeater.

Figure 1 shows the L5 apparatus case chassis, the outside-plant item required to convert an L4 apparatus case to L5. This is simply a base plate holding guide pins that line up plug-in repeaters. The guide pins serve also as retaining elements for the repeaters. Thus, the repeaters are bolted on the plate by means of the retainer guide pins, eliminating the need for the cantilevered guide and clamp arrangements used in

Fig. 2—Basic repeater apparatus case.

L4 (sometimes a source of problems because of the degree of precision required to insure proper mating of the plug-in units). The repeater is designed 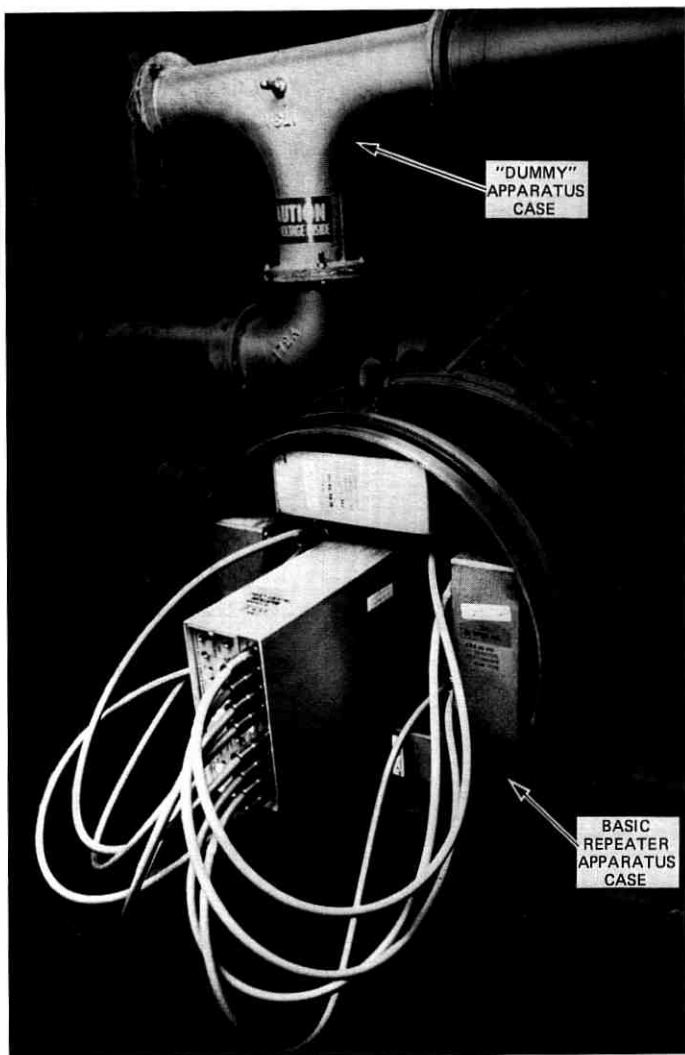so that, when bolted in place, its leading surface is in intimate contact with the apparatus case chassis. The repeater-apparatus case chassis interface becomes an important link in the thermal path from the repeater to the manhole.

Figure 2 illustrates the L5 basic repeater apparatus case arrangement featuring four basic repeaters located on the periphery of the chassis. The center space is occupied by the non-heat-producing fault-location oscillator unit. Coaxial patch cords are used for insertion of fault-location oscillator tones at both the input and output of the repeaters via twin coaxial jacks located on the apparatus case chassis. The twin jacks serve also as repeater by-pass points in the high-voltage, series-powered L5 line to permit a power patch around a repeater, without turning power down on a line, in the event repeater replacement becomes necessary.

Apparatus cases are mounted in precast concrete manholes (6 by 12 by $6\frac{1}{2}$ feet high inside the basic repeater manhole) that are not intended to be impervious to water. The apparatus cases are watertight and are filled with dry nitrogen to 9-psi cable pressure to protect against water seepage. Six apparatus cases are required in a basic repeater manhole to fully equip a 22-tube coaxial cable. Adaptors, or "dummy" apparatus cases (one of which is shown at the top of Fig. 2), are installed to permit an entire manhole to be racked and spliced initially. Apparatus cases are installed only to cover the number of tubes being equipped in order to defer expenditures until necessary.

The cross-connect apparatus case, an additional watertight apparatus case that is somewhat different in design, is also located in the manhole. It houses both the logic unit (used to turn on fault-location oscillators) and the appearances of all the coaxial cable's copper wire pairs used for support systems such as fault location, order-wire, alarm, and surveillance. A logic-unit support-shelf kit was designed to permit conversion of this type of L4 apparatus case for use on L5.

II. MANHOLE REPEATERS

2.1 Basic repeater

Figure 3 is an L5 basic repeater with covers removed. Elements of the basic repeater are the fundamental building blocks of the L5 system and appear in every repeater in the L5 equipment.[2] These basic elements, the preamplifier, the power amplifier, the low-frequency networks, and the line-build-out (LBO) network, are assembled to an inner chassis located inside a rugged die-cast aluminum outer housing. Figure 4 is a simplified repeater cross section showing the relative physical positioning of the apparatus on the inner chassis which is assembled to a flange protruding from the inner surface of the outer housing. The low-frequency networks protrude into the power separation filter (PSF) cavity through a cut-out in the flange.
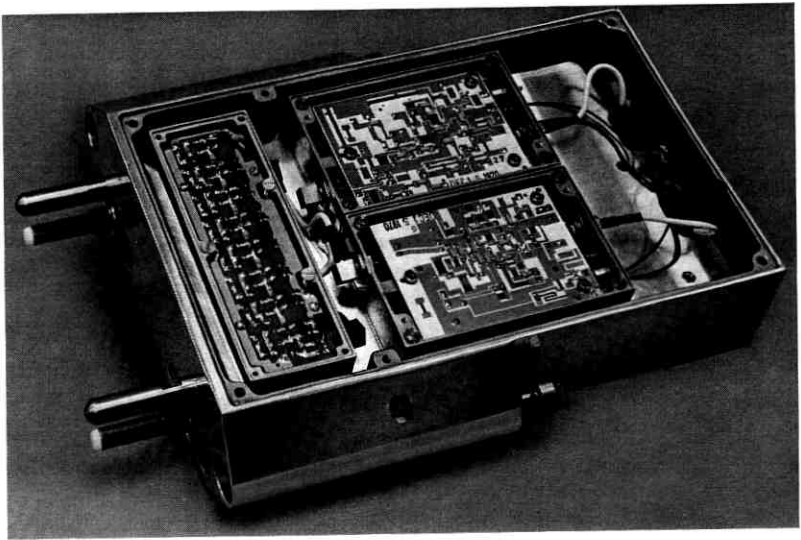
Fig. 3—Basic repeater.

Because of the series-powering arrangement for the L5 repeatered line, repeater circuit ground can be as much as ±1150 volts dc (1750 volts dc under trouble conditions) from earth or sheath ground. In Fig. 4, the cross section of the inner chassis assembly which is at circuit-ground potential is shown clear to contrast with the heavy black cross section of the outer housing, which is at earth-ground potential. Insulation between these two items is provided by a conformal coating of epoxy, 0.015-inch thick, applied to the inside cavities of the outer housing. A thin layer of epoxy sheet adhesive is used to bond the inner to the outer. The controlled-thickness epoxy interface provides the structural strength required, preserves the integrity of the insulation for the safety of personnel and the protection of precision, low-voltage electronic components, and constitutes a very low impedance to the thermal path.

The insulating and bonding techniques are the same as those developed for the L4 repeaters. As in L4, the bond joint was designed to cover a relatively large area for strength and for enhancement of thermal conduction, which is further enhanced by the massive cross sections of the repeater's structural elements. The epoxy is applied by the fluidized bed process. The process is carefully controlled to eliminate voids, thereby minimizing the likelihood of high-voltage corona "popping."

The individual apparatus units are contained in die-cast aluminum cans fastened to the inner chassis by threaded inserts integral to the inner chassis. The repeater covers are coated with epoxy on the inside to complete the continuity of the insulation of the two cavities. They are secured using one-way screws as a safety feature. In effect, "sealed" repeaters are shipped to the field. There are no field adjustments to be made on a basic repeater, and repairs, if necessary, are made at the factory. Therefore, there is never a reason to open a repeater in the field. They are sealed to avoid the exposure to high voltage that would occur if a coverless repeater were accidentally inserted into a powered apparatus case. Thus, safety features are not compromised by the use of the simplified apparatus case chassis design (Fig. 1). High voltage appears on the center conductor of the coaxial jacks that are located deep inside the apparatus case, and sealed repeaters keep the high voltage from the reach of craftspeople.

2.1.1 Hybrid-integrated circuits

The preamplifier and power amplifier are the heart of the L5 repeaters. They are thin-film hybrid integrated circuit amplifiers, and in the basic repeater they are located side by side in the center part of
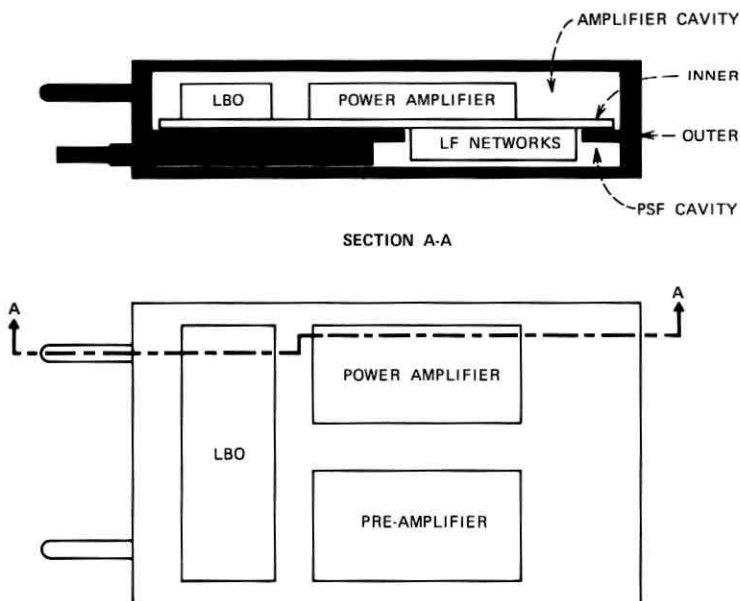
Fig. 4—Basic repeater cross section.

the repeater as shown in Fig. 3. Even though the amplifiers require many discrete components such as transformers, inductors, capacitors, and transistors, it was established very early in the L5 development that conventional printed-wiring arrangements could not be used. The precision required for L5 amplifier circuits was such that thin-film circuits were essential to the amplifiers. However, L5 required large substrates, holes in the substrate to mount discrete components, low-value precision resistors, and high-conductivity conductor paths. Each of these requirements suggested a departure from generally accepted thin-film practice. Together, they constitute a unique set of requirements and resulted in HIC's that are unique in their physical and electrical attributes.

Although most HIC's are small enough to fit many patterns on a single $3\frac{3}{4}$ by $4\frac{1}{2}$-inch ceramic mirror, the L5 HIC's are large enough ($2\frac{1}{2}$ by $3\frac{1}{2}$ inches) to allow only one per mirror. Conventional, leaded components are almost never mounted on a ceramic substrate, but this was required here. This gave rise to the need for about 50 holes per substrate. Holes complicate the manufacture of substrates and are avoided, if possible. To produce the low-value precision resistors, the resistive metal layer was deposited to a thickness of 25 ohms per square, as opposed to the conventional thickness of 50 ohms per square. This allowed longer resistors and a larger proportion of anodized-to-un-anodized area. The higher percent of the area anodized results in greater long-term stability. Furthermore, thicker metal film is more stable because resistance changes are caused by surface changes, and the surface constitutes a smaller portion of a thicker metal film than of a thinner metal film. An operating temperature of less than 90°C is necessary to maintain end-of-life tolerance for the precision resistors. Power resistors, on the other hand, would stay within their required resistance tolerance at film temperatures as high as 110°C. The power resistors and the discrete transistors are virtually the only heat producers on the substrate. Therefore, the layouts were designed so that the power resistors skirted the periphery of the substrate. The transistors, however, had to be centrally located because of circuit requirements for short connecting conductor lengths. Massive heat sinks were designed to divert the transistor heat from the precision resistors and other temperature-sensitive components.

Figure 5 is the L5 power amplifier. The top view shows the two heat sinks, each housing a matched pair of transistors, surrounded by other discrete components. The bottom view, the thin-film side of the hybrid integrated-circuit amplifier, illustrates the film topology and a number
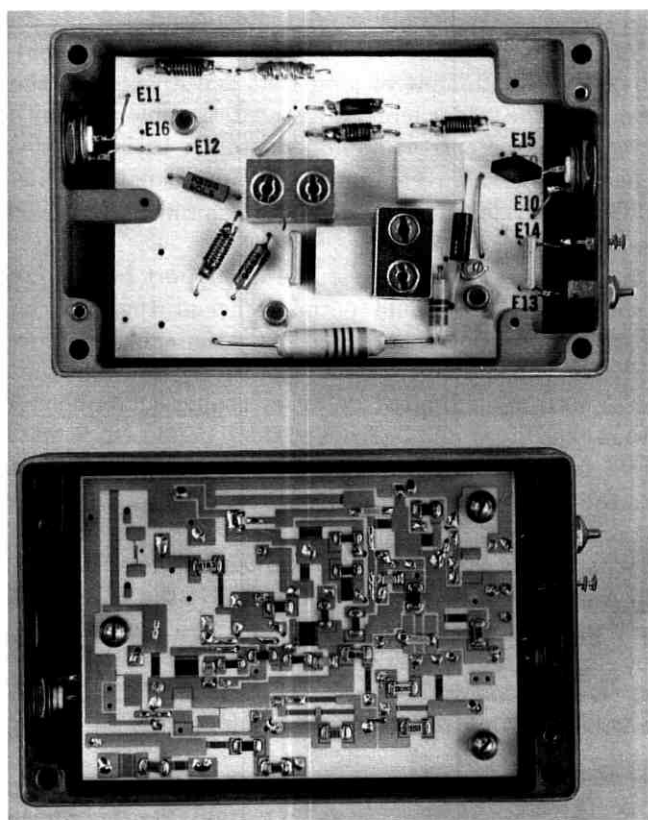
Fig. 5—Power amplifier.

of reflow-soldered barium titanate chip capacitors. A special conductor metal system was developed for L5 to produce the needed high-conductivity paths and to provide solderable pads for the chip capacitors and the leaded conventional components. The high conductivity is obtained by depositing "thick" gold. Electroplated rhodium is added to prevent the gold from dissolving in the solder. A gold flash over the rhodium protects the rhodium from oxidation, thereby retaining its solderability. By selectively etching the gold to expose a rhodium strip around the solder pads, solder "dams" are formed after the exposed rhodium strip becomes oxidized and nonsolderable when exposed to the elevated temperature of the film pre-aging process. Therefore, during the solder reflow process, when the chip components are connected into the circuit, the exposed rhodium strips confine the solder

to the area they surround. This avoids wetting and possibly degrading the high conductivity paths. By precisely defining the solder area, thereby limiting the amount of gold that can dissolve in the solder, embrittlement of the solder joints is avoided. Complementary developments determined the correct amount of solder to be supplied as an integral part of the chip component terminations and the precise reflow soldering cycle, thereby assuring reproducible, reliable HIC assemblies.

It was necessary to develop a structural support for the large, thin ceramic substrates that would permit efficient thermal conduction from the transistors, but that would keep the substrate free of mechanical stress. The substrates are not perfectly flat, and the amount of bowing or warping is limited by die-screening subsequent to firing. Ideally, the substrate would be fixed at three points (to define a plane) and contact to the transistor would be by a flexible heat conductor. However, the requirements are inconsistent for providing, simultaneously, a massive heat sink for adequate heat dissipation and a flexible heat sink for stress-free mounting. Consequently, the structural support developed was to hard-mount the substrate on the massive heat sink and merely stabilize its outer periphery on three bosses of the die-cast aluminum amplifier housing. The location of the three bosses and the stabilizing screws is shown in Fig. 5, and a partial cross section of an amplifier assembled in a repeater is shown in Fig. 6.

2.1.2 Thermal path

In Fig. 6 the large substrate, in cross section, appears virtually as a beam with a fixed support in the center. The end supports (only one shown) are designated "grommet suspension." At these points, the substrate is held between two soft, rubber grommets confined in a large hole in the substrate by a shoulder screw anchored to the boss in the amplifier housing. The role of the grommets is to stabilize the nonflat substrate in whatever position it may happen to fall and have it remain there, stress-free.

The substrate's fixed support is the massive transistor heat sink, also shown in cross section in Fig. 6. The discrete transistor, located inside the heat sink, has a beryllium-oxide header hermetically sealed to a Kovar* case. Heat sinking is to the beryllium oxide, which has the higher coefficient of thermal conductivity, and thermal flow through the transistor-heat-sink interface is enhanced by the force exerted by

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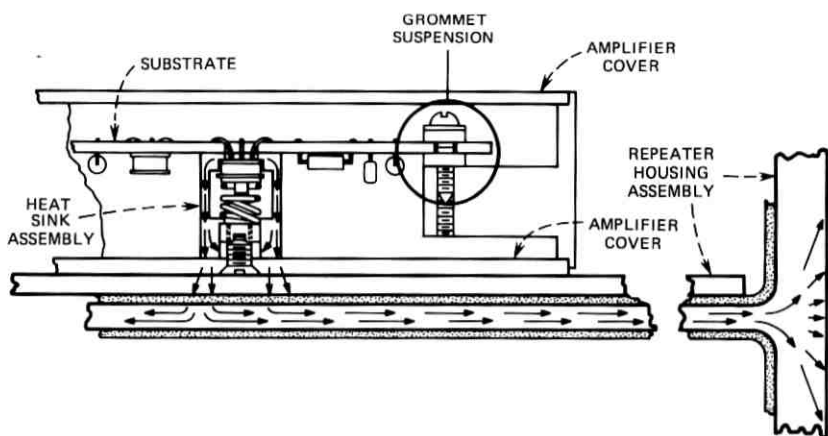* Trade name of Westinghouse Electric Corporation.

Fig. 6—Thermal path for L5 HIC amplifier.

the preloaded spring confined within the heat sink. While high, compressive pressure is applied through the transistor to the heat sink, none of it is transmitted to the delicate thin-film substrate to which the transistor is soldered. The result is a stress-free structural support for the large L5 substrates and an effective thermal circuit. As shown by the arrows in Fig. 6, heat is conducted from the transistor through the heat sink, through the amplifier cover into the inner repeater chassis, and through the thin layer of epoxy to the outer repeater housing.

2.1.3 Heat sink

Close spacing of the transistors on the substrate was required to minimize parasitic inductance and capacitance in the electrical circuit and resulted in a unique heat sink design. Spring-loading the transistor within the confines of a massive heat sink was developed for use in the L4 system amplifiers. However, the L4 scheme was not applicable to L5 because of the combination of space limitations and the application of the heat sinks to delicate, large, ceramic substrates in L5, in contrast to the rugged epoxy-glass printed wiring boards used in L4. The L5 heat sink is a precision assembly that, for good heat transfer, presents the largest possible metal cross-sectional area within the allowable limits for the transistor lead lengths and transistor spacing. The screw that preloads the spring against the transistor serves as the threaded nut for mounting the heat sink against the flat amplifier cover. To minimize thermal impedance, the heat sink's functional faces have a

32 micro-inch finish, and an indium washer is inserted at the transistor-heat-sink interface. This assures intimate contact at the thermal circuit joints.

The heat sink's cavity, large enough to accept the transistor with its flanged case, is constricted at the threaded portion because of the transistor spacing constraints. This constriction makes it necessary to have a side opening in the heat sink to load the transistor and precludes the use of conventional screw-machine methods of manufacture. Double heat sink designs are used in the preamplifier and power amplifier to maximize the amount of metal usable for heat sinking within the transistor spacing constraints. This works out well, because the transistors are used in matched pairs and can be preassembled into the heat sinks for subsequent assembly onto the substrates.

2.1.4 Individual packages

The thin-film substrates are housed in die-cast aluminum cans. These cans require a certain degree of precision to provide accurate stress-free positioning of the substrates, as shown in Fig. 6. Substrate supporting surfaces are held within tight tolerances with respect to the cover supporting surface. The bottom amplifier cover is controlled for flatness and is designed to be overflush to present maximum contact area to the repeater inner chassis for heat transfer.

Having top and bottom amplifier covers removable provides accessibility to both sides of the substrate during shop processing and also for trouble shooting, and the open die-cast frame of the amplifier housing serves as a convenient carrier and substrate protector during handling in the shop.

Connections to the substrate are made through 75-ohm miniature coaxial connectors that clearly define electrical and mechanical boundaries for the amplifiers. This substantially facilitates amplifier testing, repeater assembling, and, if needed, repair and replacement.

2.1.5 Maximum transistor junction temperature

To ensure reliable performance of silicon solid-state devices, the physical design goal was to limit the maximum junction temperature to 125°C and, in the case of the elements of the L5 repeatered line, can be expressed as

$$T_J \, \text{MAX} = \Delta T_{AC} + \Delta T_R + \Delta T_T + T_M \leq 125°C, \qquad (1)$$

where

T_J = transistor junction temperature,
ΔT_{AC} = temperature rise in the repeater apparatus case,
ΔT_R = temperature rise in the repeater,
ΔT_T = temperature rise from case to junction of the transistor because of its thermal impedance,
T_M = maximum manhole ambient temperature.

For the L5 basic repeater, T_M has been calculated at 37°C for the hottest manhole in the hottest part of the country, based on an empirical expression relating manhole power dissipation, soil condition, and manhole dimensions. ΔT_T for a device is fixed for its specific power dissipation and heat sinking environment, and the design features to control ΔT_{AC} and ΔT_R were described earlier in this paper.

The maximum junction temperature for transistors in the L5 basic repeater power amplifier has been determined by tests to be 128°C, and comparative performance of L4 and L5 is shown in Fig. 7. Recalling that an L5 repeater dissipates some 18 watts in comparison to $13\frac{1}{2}$ watts for L4, it is evident that the thermal conductivity enhancement was realized as anticipated because of the apparatus case chassis design and the repeater packaging techniques developed for L5. At 128°C, maximum transistor junction temperature for L5 transistors, thermal results of L4 and L5 are equivalent and, based on L4 performance to date in the field, it appears that high-reliability performance of L5 devices is ensured.
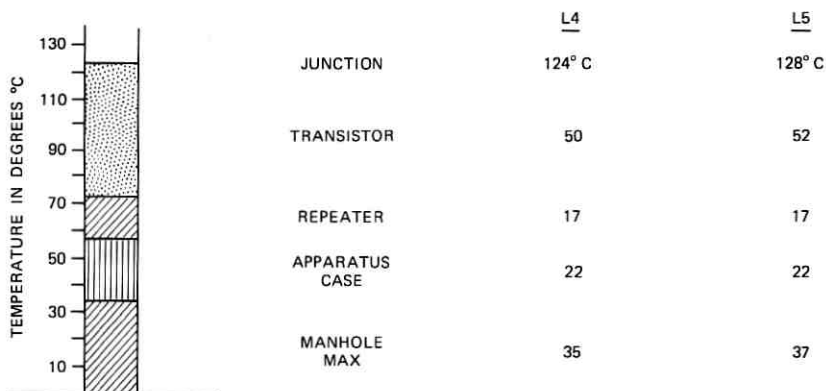
		L4	L5
	JUNCTION	124° C	128° C
	TRANSISTOR	50	52
	REPEATER	17	17
	APPARATUS CASE	22	22
	MANHOLE MAX	35	37

Fig. 7—Maximum transistor junction temperature.

2.2 Regulating repeater

The "modified-conversion" design concept as applied to the L5 regulating repeater requires that two repeaters fit in one apparatus case. The two repeaters are bolted to a newly designed apparatus case chassis, similar to that for the basic repeater illustrated in Fig. 1, in a layout that positions the fault-location oscillator unit between the two repeaters. A maximum of 11 apparatus cases are required to accommodate an L5 22-tube coaxial cable, and these cases are arranged on one wall of a precast concrete manhole 6 by 12 by 9 feet high on the inside, $2\frac{1}{2}$ feet higher than a basic repeater manhole.

The regulating repeater package is one of the most difficult of the L5 physical designs. A number of design constraints were imposed, such as field access for adjustments, addition of constituent apparatus and in-service testing, heat dissipation, high-voltage insulation, and personnel safety.

This repeater contains the same circuitry as a basic repeater with the addition of regulating circuitry, space for field insertion of two line-build-out networks and a deviation equalizer, and a low-voltage test point for in-service measurements. Like the basic repeater, a rugged, epoxy-insulated outer aluminum die-cast housing holds the captive retainer screws for bolting the repeater to the apparatus case chassis. Unlike the basic repeater which has a flat-plate inner chassis, the inner chassis of the regulating repeater is compartmented to accept individually packaged circuit sections and units that do not require their own individual shield cans (see Fig. 8). Because of the packaging density required to have two repeaters cantilevered from the apparatus case chassis in a single apparatus case, the center flange of the outer housing had to be kept to a minimum, allowing just sufficient area to ensure a good bond to the inner chassis. Therefore, compartmenting, in combination with individual apparatus housings that act as stiffeners, produces an array of ribbing on both sides of the large inner chassis and provides the rigidity required over the large cross-sectional area of this repeater, which measures $10\frac{1}{2}$ inches wide by $17\frac{1}{2}$ inches long.

2.2.1 Hybrid-integrated circuits

The regulating repeater uses five thin-film hybrid-integrated circuit (HIC) amplifiers in addition to the preamplifier and power amplifier. The five are flat-gain amplifiers, and their ceramic substrate size is more nearly conventional, measuring 0.8 inch by 2.9 inches. Actually, three designs, two of which appear twice as part of the two $\sqrt{f}$ net-
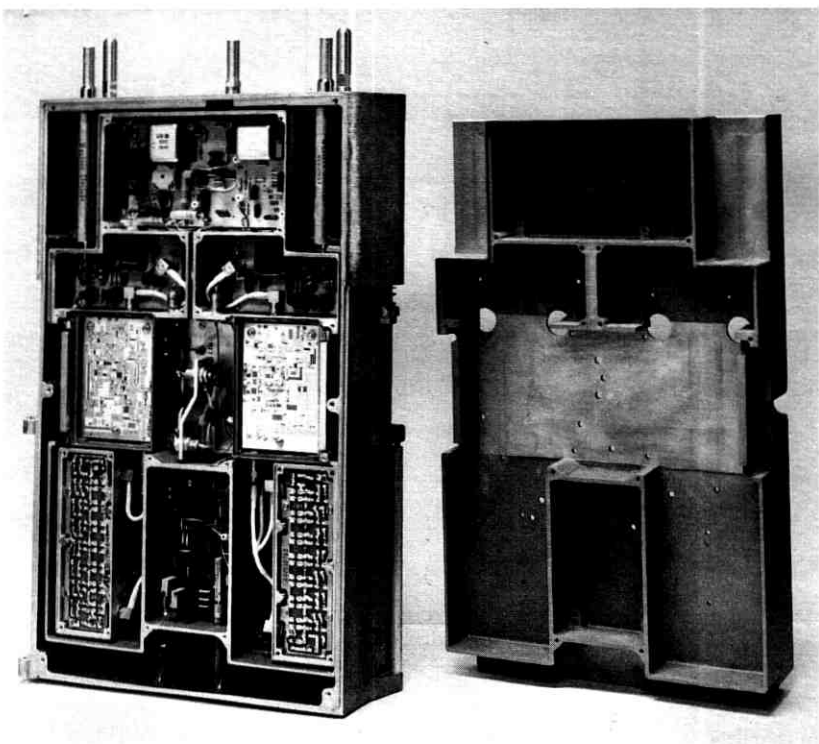
Fig. 8—Regulating repeater.

works,[3] are used in the regulating repeater. In the $\sqrt{f}$ network, the HIC's are connected as slave-boards to epoxy-glass printed-wiring network boards by means of 26-gauge leads on the HIC's. They are formed to fit into holes in the printed-wiring board and are clinched to touch land areas to which they will be subsequently soldered. The substrates are structurally supported on transistor heat sinks that enclose the discrete, heat-producing transistors and are fastened directly to the can that houses the entire network. The heat sinks are similar to those described for the preamplifier and power amplifier, except that they are the single-cavity style. Here, too, conduction is the primary mode of heat transfer. The heat from the transistors is directed away from the substrate, through the heat sink into the die-cast network housing and to the repeater framework, etc. Top and bottom network covers are removable to provide accessibility
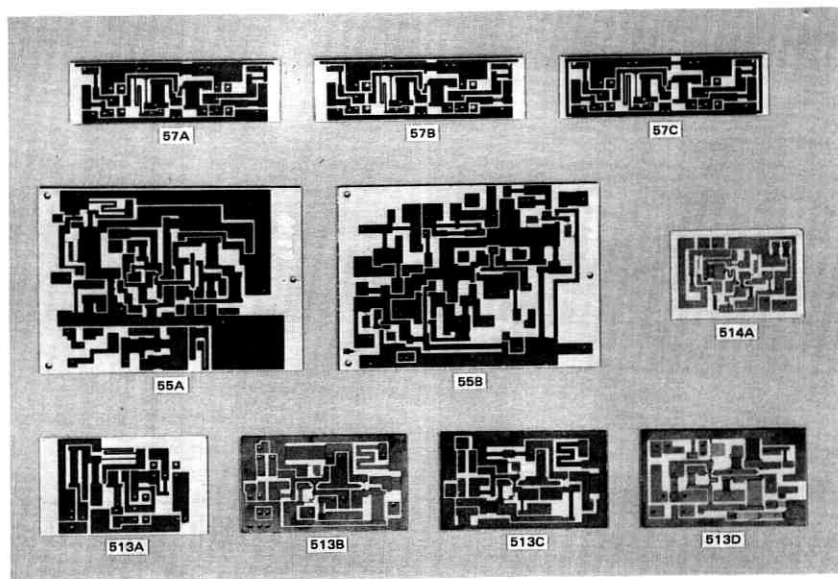
Fig. 9—FIC's used on L5.

for trouble shooting but, more important, to allow assembly in a prescribed sequence. The printed-wiring board is first mounted in its frame, then the HIC is fastened to the frame by the heat sink and finally the HIC leads are soldered to the board land areas. This assembly sequence is essential to a stress-free assembly. Figure 9 shows the collection of thin-film integrated-circuit substrates used throughout L5. The 55-types and the 57-types are the ones described earlier for use in the repeaters, while the remaining ones are used in flat-gain amplifiers for terminal equipment described later in this paper.

2.2.2 Buried thermistors

To compensate in part for changes in cable loss because of changes in cable temperature,[2] the regulating repeater requires access to ground-temperature-sensing thermistors that are buried 15 feet from the manhole at cable depth. The thermistors are contained in a cast-epoxy cable stub and, at a distance of 15 feet, are unaffected by manhole temperature. A plug-and-jack connection through the apparatus case chassis provides the thermistor connection to the repeater.

2.2.3 Safety

While the basic repeater is shipped to the field as a sealed unit, the inside of the regulating repeater must be accessible for the addition of line-build-out networks and for adjustments when the repeater is acceptance-tested in the field.[2] For this reason, the regulating repeater's two large covers are hinged as a safety feature to prevent exposure to high voltage by the accidental insertion of a coverless repeater into a live apparatus case. Manipulation of the bulky, hinged covers is awkward, at best, during manufacture and field-acceptance testing, but amounts to a relatively effective trade-off for a necessary safety precaution.

High-voltage insulation lines the cavity on the inner chassis that houses the coaxial jack used as the low-voltage access point for the repeater. Even though the jack is electrically isolated from high voltage by a transformer and is mechanically bonded to the outer housing which is at earth or sheath ground, the insulation liner is provided as a safety feature to protect personnel and equipment in the event a fault should occur.

2.3 Equalizing repeater

With equalizing repeater points located at 34-mile (maximum) intervals, or roughly midway in the 75-mile (maximum) L5 power spans, it is unlikely that, in conversion from L4 to L5, they would coincide with L4 equalizing repeater points spaced at 54-mile (maximum) intervals.[4] Therefore, the only constraint on the L5 equalizing repeater, to comply with the modified-conversion design concept for L5, is that similar apparatus cases be used for the sake of standardization.

However, L5 uses manually adjustable equalizers with less circuitry than is required for the remotely adjustable equalizer used in L4. This simplifies manhole arrangements in that the entire L5 equalizing repeater was designed to fit in a single apparatus case, in contrast with the two cases required to house the L4 equalizing repeater. For a 22-tube coaxial cable, an equalizing repeater pre-cast manhole measures 6 feet wide by 24 feet long by 8 feet high on the inside to accommodate the layout for the 22 apparatus cases required to equip the entire cable.

The equalizing repeater comprises three plug-in units—a regulating repeater, a manhole E1 equalizer (ME1), and a fault-location oscillator unit arranged in that order from left to right in the apparatus case. As with the previous repeaters, these units are bolted to the apparatus
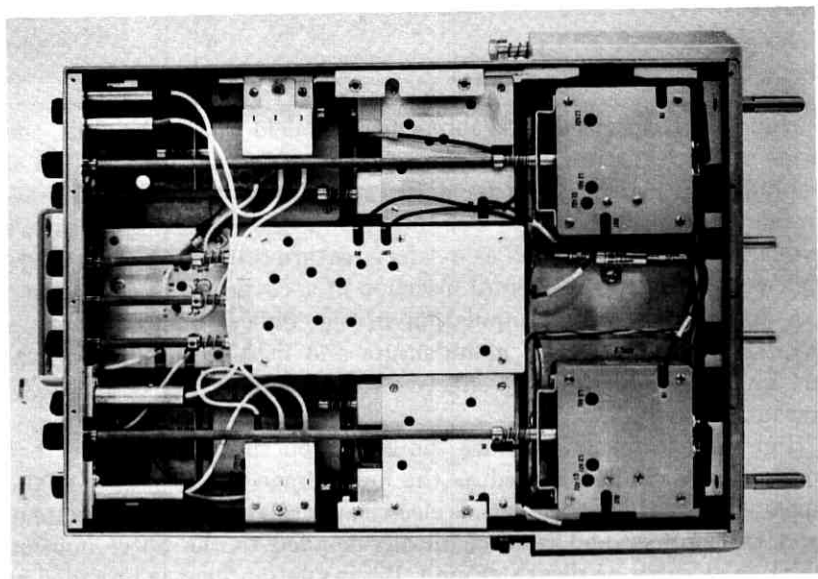
Fig. 10—Manhole E1 equalizer.

case chassis and are cantilevered inside the apparatus case. Equalizing repeater points also provide access to buried-cable-temperature-sensing thermistors.

The regulating-repeater portion of the equalizing repeater is identical to the regulating repeater described earlier except that one low-voltage test-access jack is replaced by two low-voltage jacks for connection to the ME1 equalizer.

The ME1 equalizer is a larger unit than the regulating repeater, and was designed as a fabricated unit to keep its weight down. The unit measures 12 inches wide by 17 inches long by $4\frac{1}{2}$ inches deep, and is illustrated in Fig. 10. The array of individually packaged amplifiers and Bode networks is visible. These use printed-wiring technology, are interconnected by cables terminated in miniature coaxial connectors, and are made accessible to the knobs of the ME1 faceplate by insulated shafts that permit manual control of potentiometers for equalizer adjustments. The knobs are "push-to-turn" to prevent inadvertent changes in adjustments, and universal couplings prevent lateral forces from being transmitted to the delicate potentiometer shafts. A power separation filter is contained in the lower cavity of the ME1 equalizer, and high-voltage insulation is provided by phenolic strips. This unit, like the basic repeater, need not be opened in the

field and, for safety, it is sealed by having its covers secured with one-way screws. The mounting base of the ME1 is a one-piece aluminum casting that permits the precise location of the mounting holes and the coaxial plugs within the tolerances allowable for mating with the apparatus case chassis.

III. MAIN-STATION EQUIPMENT

Main stations in an L5 "backbone" route are single-story buildings containing four general categories of main-station equipment—high-frequency repeatered line, signal processing terminal, transmission surveillance, and a fourth, broad or peripheral category comprising such equipment as order wire, restoration, repeater acceptance test, and equalizer adjustment. The main-station equipment, intended for the central office environment, is bay-mounted in contrast to the rugged, cast plug-in type equipment designed for the manhole environment. The main-station equipment is generally in shelves, drawers, or panels in unequal-flange duct-type bays, on a 2-inch mounting plate modular spacing, and forming a bay-front that is flush with the bay uprights. The equipment space is 15 inches deep, the front and rear guard rails are 2 inches and $\frac{3}{8}$ inch, respectively, and the duct formed by the uprights of two adjacent bays constitutes shielded cabling space closed at the front and accessible from the wiring aisle at the rear of the bays through the opening created by the smaller flange (see Fig. 11). The space directly behind the cabling duct is available for shelf interconnections, for overflow cabling if required, and, in the case of a double bay, for additional equipment.

The deep equipment space is required because of the relatively bulky, well-shielded apparatus assemblies used throughout the system.
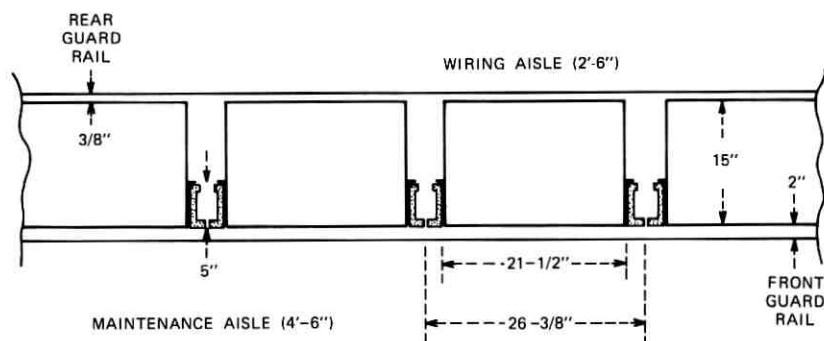

Fig. 11—Floor plan layout.

This depth, in combination with the width of the standard bay, gives rise to three-dimensional, high-density, equipment packaging which saves coveted floor space in a main station but limits accessibility of individual constituents of bay subassemblies. However, the limited accessibility is tempered by the connectorized assemblies used throughout the system to facilitate manufacture and maintenance.

Miniature coaxial connectors terminate individual pieces of apparatus (amplifiers, filters, networks, etc.) to form precise, definable boundaries for specific electrical requirements. These individual units are arranged on a shelf and interconnected by miniature coaxial cables terminated in the mating coaxial connectors. The shelf is screw-fastened to the rear flange of the bay uprights, and is itself interconnected with small Bell System coaxial plugs and jacks to the bay wiring harness for signal connections and with multicontact connectors for power and alarm connections.

In some instances, plug-in modular printed wiring board (PWB) assemblies are arranged on a shelf and interconnected by gold-plated fingers on the PWB and multicontact connectors that are part of the back-plane wiring of the shelf.

High-density packaging tends to accentuate thermal considerations, and limited accessibility coupled with high system capacity accentuate reliability considerations, both prime considerations in the design of L5 equipment.

The L5 bay height was standardized at 9 feet to accommodate power-feed main and switching power-feed main-station buildings. Terminal and terminal-main stations have higher ceilings and require bays 10 feet, 6 inches and 11 feet, 6 inches high. For these, bay extenders are added to have the 9-foot standard bay fit in 10-foot, 6-inch and 11-foot, 6-inch bay line-ups. Specific bays, intended for use only in terminal and terminal-main stations and requiring the additional space afforded by the higher ceilings, are 10-foot, 6-inch and 11-foot, 6-inch designs. The new equipment building system (NEBS) requires that future equipment be standardized on a 7-foot bay height, and a series of 7-foot bays have been designed for L5.

3.1 Main-station line equipment

3.1.1 Power separation filter cabinet

For the L5 high-frequency repeatered line, the interface between manhole repeaters and main-station equipment* is the power-separation filter (PSF) cabinet (see Fig. 12). This cabinet confines all the
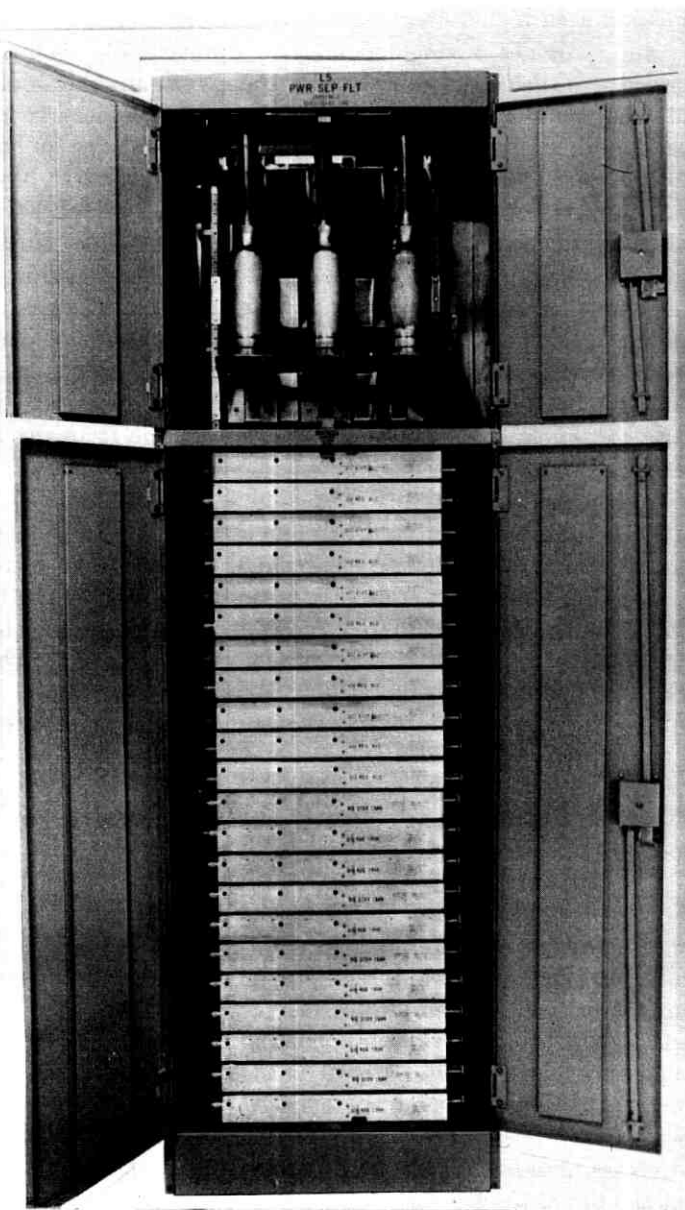
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*See Fig. 4 of Ref. 4.

Fig. 12—Power separation filter cabinet.

transmission area high-voltage equipment in one locked enclosure for safety, and provides a single point to terminate the lead-covered, pressurized, air-dielectric cable stubs coming from the multitube, coaxial-cable splice in the station. This obviates the need to run the lead cable and the high voltage in front of the line bays, as has been done in the past. The run from the PSF cabinet to the line bays is at low voltage and uses single-unit air-dielectric and/or solid-dielectric coaxial cable, depending on the length of the connection. Initially, it was intended that the PSF cabinet be located in close proximity to the building cable entrance to minimize the run of lead cable in the station. However, placing the PSF cabinet at the recommended location at the head-end of an L5 bay line-up optimizes the interbay cabling for that line-up and, in most cases, precludes the need for the air-dielectric coaxial-cable-run portion of the connection from PSF cabinet to the line bays.

The 9-foot-high PSF cabinet can hold a maximum of 22 power separation filter shelves required to equip an entire 22-tube cable. The shelves are contained in the lower portion of the cabinet seen from the rear in Fig. 12. The cable terminals are bolted to brackets in the top portion. Connections are made by screw-type coaxial connectors for the high-voltage runs and by plug-and-jack connections for the low-voltage runs. The cabling is dressed in the ducts formed on the sides of the cabinet by the standard bay uprights (used to structure the cabinet) and the fabricated metal enclosure is assembled around the bay. Bay extenders accommodate the 9-foot cabinet in higher than 9-foot equipment line-ups. The PSF shelves are modular and can be equipped on an as-needed basis.

3.1.2 Line-protection switching system

The line-protection switching system (LPSS-3)[5] bay is the second element in an L5 line-up, and is required at each end of an L5 switching section (150 miles maximum), but not at the power-feed main stations located at the 75-mile (maximum) points. This 9-foot bay contains a preponderance of logic-type circuitry that is packaged in modular fashion on plug-in PWB assemblies arranged on fixed equipment shelves in the bay. The myriad interconnecting wires are formed into a sizable cable harness, dressed in the cable duct, and connected to a terminal strip at the top of the bay. Bay power is derived from plug-in dc-to-dc converters located and fused at the top of the bay. Status indications and controls are concentrated in a large display panel located centrally on the bay. Human-engineering design considerations predominantly

influenced both the physical arrangement of this panel and the selection of the lighted indicators and keys so that the system's switching state can be clearly communicated to the craftsperson. The presence of plastic hinged protective covers on the more important controls also provide the "keying" necessary to avoid inadvertent manipulation of controls so as not to jeopardize service. Wire verification to ensure the presence of all intrabay connections is required before shipment of a bay containing the common equipment. Modular plug-in assemblies can be shipped separately on an as-required basis as the system grows.

3.1.3 Line transmit-receive bay

The line transmit-receive bays are the mainstay of the L5 line-up. One line bay is required for each transmit-receive pair of coaxials equipped in the system. This bay is essentially an equalizing repeater with manually adjustable equalizer units, E1 and E2, in both the transmit and receive sides and with the addition of a dynamic equalizer, E3, in the receive side. This, together with the elements of line-connecting, powering, and alarm equipment, comprises the line transmit-receive bay which appears at every main-station location. Four designs accommodate the 9-foot requirements at power-feed and switching power-feed main stations and the 10-foot 6-inch and 11-foot, 6-inch requirements at terminal and terminal-main stations. The difference among the four designs is the varying complexity of the elements of line connecting, ranging from the very simplest in the power-feed main-station design, where the transmit-receive bay is essentially a one-way repeater, to the more complex one-way repeater of the switching power-feed main-station design, which includes switches and switch-initiator circuitry for LPSS, to the most complex terminal and terminal-main station designs where elements of line connecting also include blocking, adding, dropping, and branching circuitry for the signal-processing features. This hierarchy of line bay designs is illustrated in Fig. 13, with shading used to contrast the variable element within the otherwise repetitive structures. Note that the only difference between the two "terminal or terminal-main" designs is that the 11-foot, 6-inch bay has room for one additional line-branching unit.

The main-station transmitting and receiving repeaters together equate to a regulating repeater used in manholes. The transmit and receive elements have been split and repackaged in an equipment shelf for bay-mounting (Fig. 14). Similarly, the main-station manually adjustable equalizer (E1) contains the same elements as its plug-in unit, manhole counterpart (Fig. 10), re-packaged in a fixed shelf for

Fig. 13—Line transmit-receive bays.
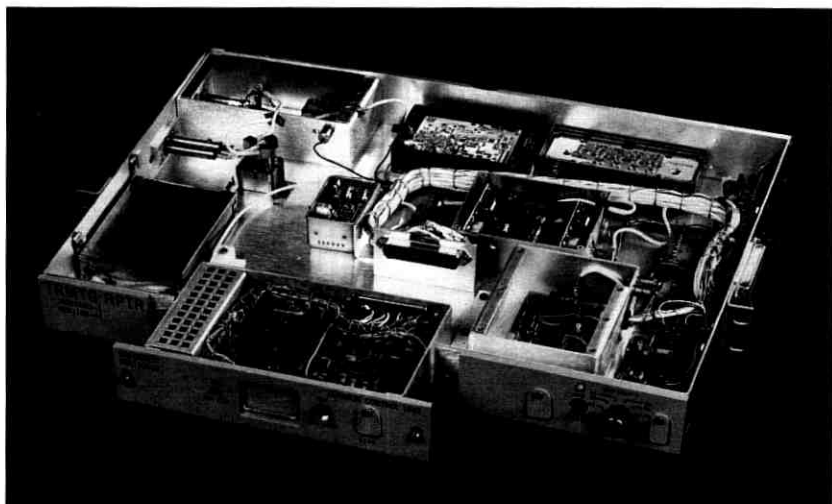
Fig. 14—Main-station transmitting repeater.

bay-mounting (see Fig. 15); and this unit, expanded to include additional networks and amplifiers required for additional bump shapes, yields the design of E2 (see Fig. 16).

Two line bays are shown at the right in Fig. 17 as they appear at a terminal or terminal-main station in an L5 five-bay line-up that constitutes the equipment required to provide service on the first working

Fig. 15—Main-station E1 equalizer.

PHYSICAL DESIGN   2171

Fig. 16—Main-station E2 equalizer.

pair of tubes, with the first line bay being the protection bay and the second the first regular bay.

3.2 Terminal equipment

3.2.1 Jumbogroup multiplex bay

The final step of multiplexing required to stack jumbogroup (JG) signals for transmission over the L5 coaxial line is accomplished by the jumbogroup multiplex[6] (JMX). The JMX is a completely solid-state multiplex that utilizes thin-film hybrid-integrated circuit (HIC) elements contained in m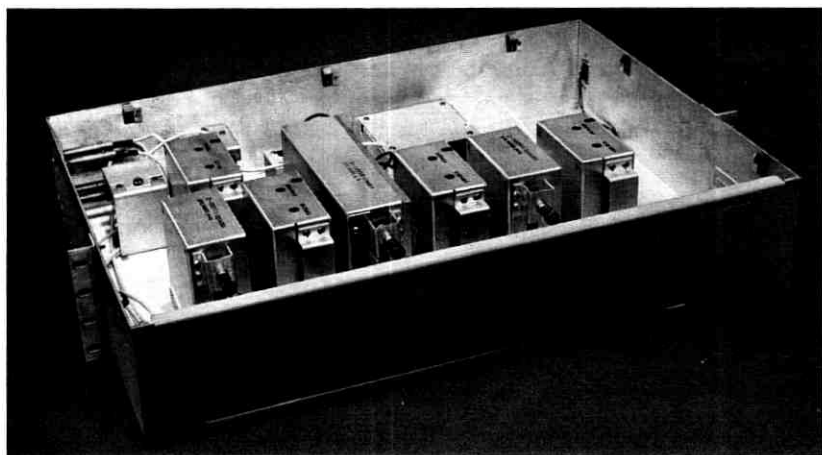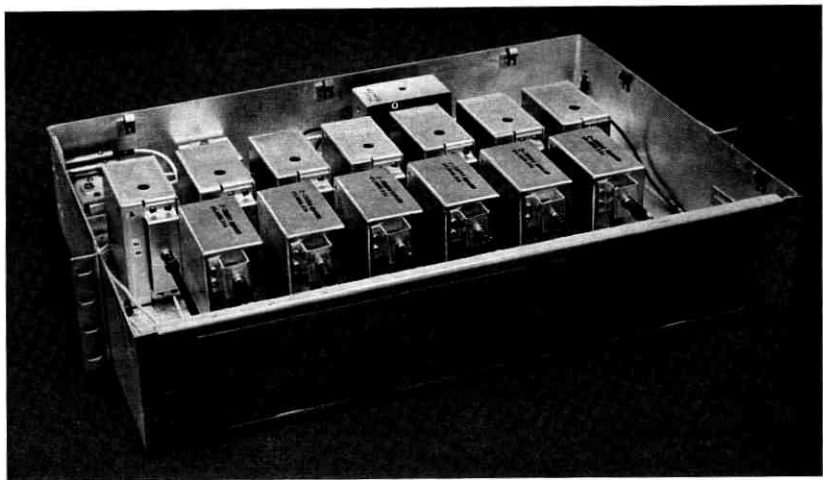odular plug-in assemblies. The JMX equipment is mounted on a unitized bay framework shown in Fig. 18, 11 feet, 6 inches high by 52 inches wide by 15 inches deep. A complete bay accommodates a maximum of three transmitting and receiving jumbogroups. Jumbogroup multiplex designs are also available in 10-foot 6-inch, 9-foot, and 7-foot bay heights to meet the needs of telephone offices with lower ceilings.

The JMX is a completely shop-assembled, shop-wired, and shop-tested bay. It contains transmission, shaping, regulating, patching, logic, alarm, carrier supply, and dc-to-dc converter equipment and, since 3600 voice circuits could be affected by a service interruption, automatic protection switching is also provided for each jumbogroup.

Jumbogroup multiplex bays are generally located in the maintenance aisle of a central office or main station where a minimum aisle spacing

of 4 feet, 6 inches is required to permit the use of rolling test consoles without restricting the operating personnel. Since this area is a center of office activity, a good appearance, without significantly increasing the cost, was considered a design objective.

The bay cabling consisting mainly of office-type coaxial cable is carefully segregated, primarily to minimize crosstalk and secondarily to facilitate installation and shop wiring. Transmitting cables are contained in the left-hand cable duct and receiving in the right-hand duct,
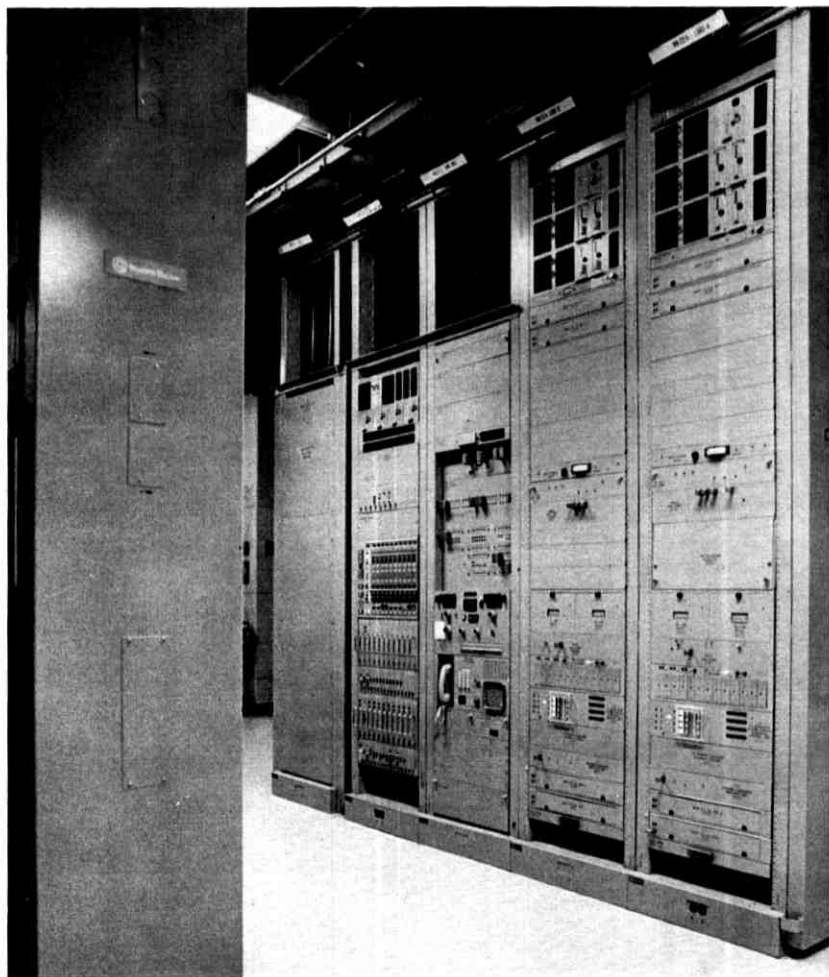
Fig. 17—L5 bay line-up.
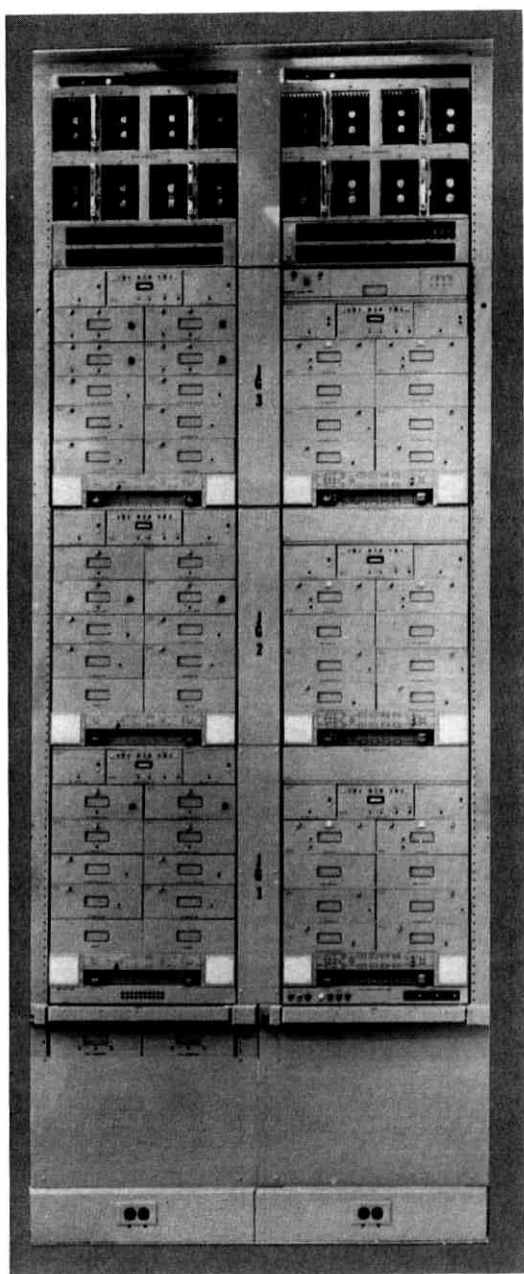
Fig. 18—Jumbogroup multiplex bay.

and, with few exceptions, all other wiring and cabling is confined to the center duct. Office cable connections to the bay are made through connectorized cables located at the top of the bay, making it unnecessary for the installer to enter the cable ducts.

The basic building block of the JMX bay is the jumbogroup shelf assembly shown in Fig. 19, which contains the modular plug-in units. For example, the receiving shelf contains the regulator, equalizer, demodulator, de-emphasis, transmission protection switch, logic switch control, and jack field. The framework and module construction for the transmitting side is the same.

Since all equipment modules performing similar functions are the same physical size, a variety of bay configurations can easily be accommodated. For instance, at a typical end office, the JMX bay with jumbogroups 1, 2, and 3 could be provided, whereas at intermediate stations the bay could contain all jumbogroups 1, 2, or 3 or any combination of jumbogroups 1, 2, or 3 up to the bay capacity of three.

Fig. 19—Jumbogroup shelf assembly.

A lattice-type lightweight aluminum construction is used for the jumbogroup shelf to optimize the strength-to-weight ratio and virtually eliminate shelf deflection under load. To avoid scraping when the aluminum modules are inserted in or removed from the shelf, a low-friction tape is bonded to the load-bearing surface.

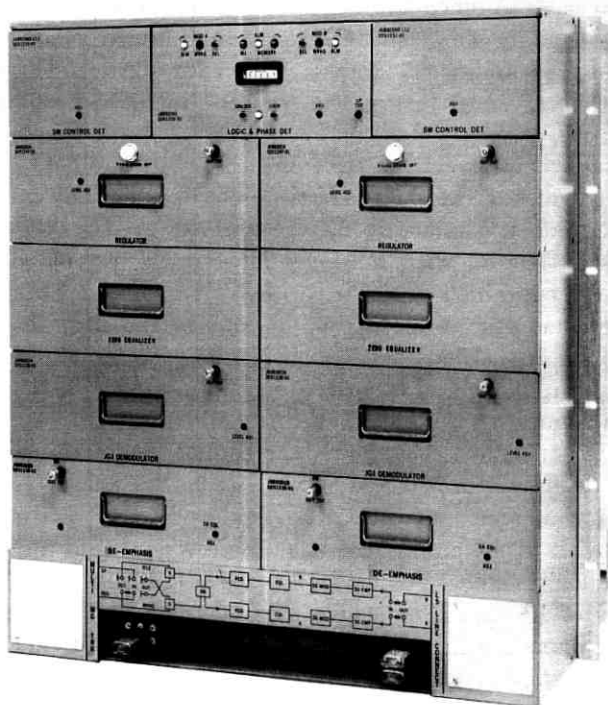Each transmitting and receiving shelf assembly contains a built-in jack field for jumbogroup patching or testing. To facilitate maintenance, a functional schematic diagram is provided directly above the actual jack location. The jack field is recessed to minimize the possibility of a circuit patch plug accidentally being removed or damaged, thereby causing a service interruption. To aid the craftsperson in identifying jumbogroup equipment with its associated jack field, each jumbogroup position is framed with a colored plastic strip. Each of the three jumbogroup bay positions is identified with a different color. Matching color route assignment cards covered with glare-resistant plastic are also provided.

Directly above the writing shelves is a miscellaneous panel that provides access to office trunks, alarm cutoff switches, and indicators and the means for simultaneously testing all lamps in the bay. In the present bay, incandescent lamps are used for alarm indications; however, in the very near future, these will be replaced by light-emitting diodes (LED).

Modules are inserted from the front of the bay and held by a captive screw in the rear of the shelf. These units are essentially plug-in, without fixed mating connectors on the shelves. In this way, tolerance problems associated with mating connectors are avoided. Most assemblies are contained in 10-inch wide by 4-inch high by 12-inch deep aluminum modules, as shown in Fig. 20. Coaxial and pin connectors are provided at the rear for transmission and power connections.

Filters, amplifiers, and other types of apparatus have been mounted in separate housings for shielding, to facilitate manufacture and testing, and for efficient field repair. Interconnections within the modules are made by miniature coaxial cables.

During the development of the JMX and other L5 main-station equipment, it became apparent that there was a need for a family of very small amplifiers. To meet this need, the 509-, 510-, 511-, and 512-series of HIC amplifier codes were developed. More than 230 of these amplifiers are in a fully equipped JMX bay. The outside dimensions of the amplifiers are 2.75 by 1.5 by 2 inches high. Input and output connections are made through miniature 75-ohm coaxial connectors. The circuitry of the amplifiers is a HIC with some components
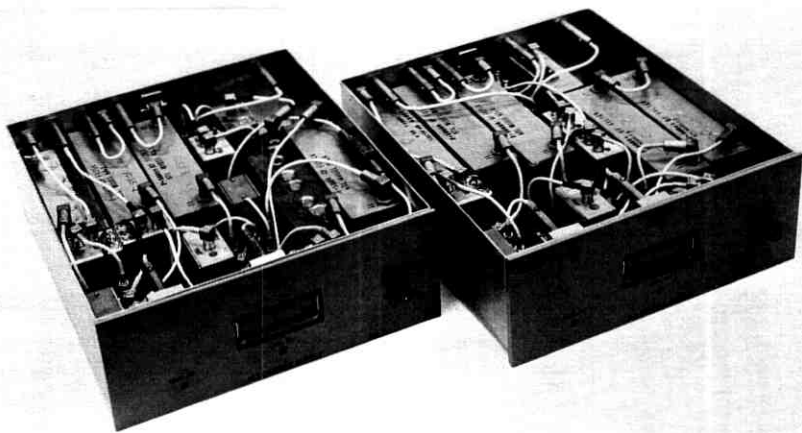
Fig. 20—JMX modules.

appliqued, such as transistors, inductors, or chip-capacitors. Special ground terminals are swaged into the chassis to ensure effective grounding.

Considerable care was taken to limit the mechanical stresses between the ceramic substrate and the chassis. For example, all connections are made with soft gold-coated copper ribbon wire, looped to allow slight mechanical movement of the substrate during thermal and mechanical shock. As shown in Fig. 21, the substrate is mounted by fitting the ends with silicone rubber gaskets and supporting them at opposite ends of the chassis with adjustable aluminum blocks. To efficiently conduct heat from the transistor, a spring-loaded heat sink described earlier connects the transistor mounted on the substrate directly to the cover of the amplifier.

As part of the overall JMX development, a complete thermal analysis of the bay was performed. Where possible, units dissipating the most power were arranged in the bay to avoid hot spots. A 4-inch space is provided for each jumbogroup shelf through the recessed jack field for a front-to-rear air flow. Temperature measurements indicate a 55°F maximum gradient between a transistor case in the JMX and room temperature. The complete thermal analysis shows that, even under the worst condition, the equipment in the bay is well within the temperature limits necessary for reliable performance.

The 24-volt power for the JMX is supplied by the office power plant over two separate feeders to the 24-volt to 25-volt regulated converters.

HEAT SINK   MTG BLOCK   RUBBER GASKETS

LEADED COMPONENT SIDE   FILM CIRCUIT SIDE
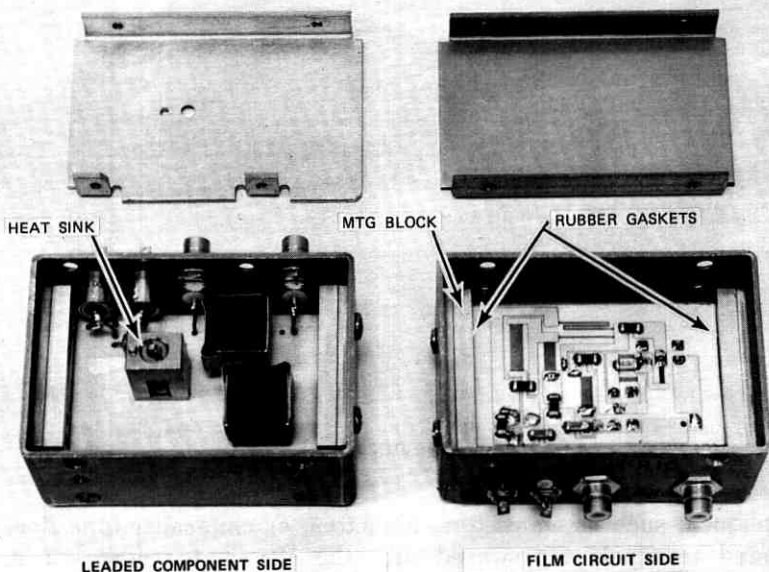
Fig. 21—HIC amplifier.

The feeder-converter assignments have been arranged so that the loss of a feeder or a single converter will not result in a loss of service.

3.2.2 Basic jumbogroup trunk bay

The interface between JMX and the lesser units in the multiplex hierarchy is the basic jumbogroup trunk bay[7] (BJGT). This bay offers a very high degree of flexibility for the user in that it is designed to allow a maximum of four inputs per jumbogroup and to include designs for inputting radio signals, L-carrier mastergroup digital (LMD) signals, L4 coaxial-carrier system signals, etc. The active circuits are redundant and are packaged in pull-out drawers and fixed shelves. The combinations of circuits available are so numerous that normal bay coding or even normal panel coding is virtually impossible. Only typical bay layouts are offered as a guide for ordering, and the specific short- and long-range requirements of an office dictate the composition of any BJGT shipped from the factory. Intra-bay cabling is laid out to accommodate multi-input options and suggested bay layouts allow space for modular future growth.

3.2.3 Jumbogroup frequency supply

The jumbogroup frequency supply (JFS) furnishes three highly reliable and precise signals to the JMX to generate the necessary carriers.[8] These signals, at frequencies of 1.024, 2.560, and 20.48 MHz, have an accuracy well within the one part in 10^8 requirement imposed by the JMX. This accuracy is realized by comparison with a reference signal from a primary frequency source; however, even if the reference signal is lost for several weeks, the inherent stability of the JFS will maintain the required accuracy.

The JFS, shown in Fig. 22, is mounted in a unitized double 9-foot high framework. A front aisle is required for normal maintenance and a rear aisle for installation wiring. Since the number of JFS bays required for coaxial systems is relatively low, the 9-foot bay arrangement is provided for use in current design of main stations. A 7-foot version of JFS is also available for future offices with the lower ceiling height.

The JFS is a fully shop-assembled, shop-wired, and shop-tested bay. It also is completely solid-state, utilizing digital and thin-film hybrid-integrated circuit (HIC) technology. The most important design objectives were precision and high reliability. To accomplish this, very stable oven-controlled 39-type oscillators and one-for-one redundancy automatically switched are employed.

The jumbogroup frequency generator (JFG) is the basic building block of the JFS bay, and the 39-type oscillators are the heart of the JFG. Each JFS contains three JFG's (master, regular, and standby). The master JFG serves as a buffer between the reference signal and both the regular and standby JFG's, thereby keeping the JFS output relatively immune from loss and/or disturbance of the reference signal. As shown in Fig. 23, an integral part of the JFG is the alarm display panel, which indicates whether the JFG is operating as the master, in-service, or idle. It also indicates frequency offset, level failure, and the frequency alarms. Test access points are provided for measuring the 25-volt, 6-volt, and 5-volt power.

The JFG patch panel provides patching facilities not only to remove a failed JFG from service but to rearrange the JFG's by function (master, regular, standby) to realize the best use of the remaining two. When a JFG is removed from service, the protection switching logic must be restructured to accommodate rearrangement of JFG functions. This is accomplished by operating the JFG "out-of-service" switch on the alarm and switch control panel when the patching has been completed. The faceplate of the JFG patch panel is a schematic block diagram with
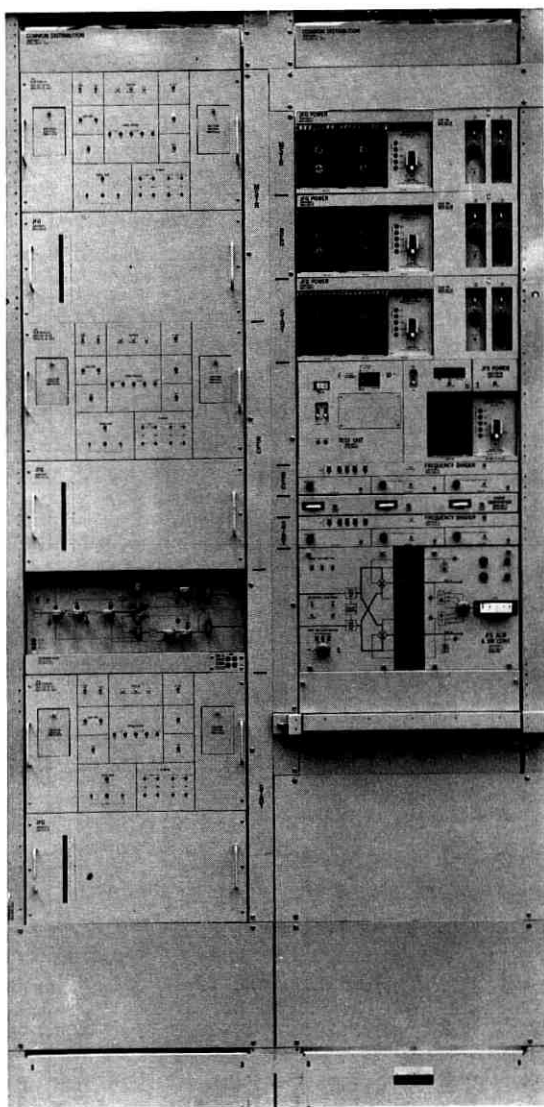
Fig. 22—Jumbogroup frequency supply bay.

appropriately positioned jacks interconnected with color-coded paths to indicate the placement of patch plugs for the JFG, master, regular, or standby failure conditions. As in the JMX, the panel is recessed to prevent damage and possible circuit interruption.

The control center for the JFS is the alarm and switch control panel. It is a visual continuation of the patch panel schematic block diagram and contains the level detectors and coaxial switch and the logic required for the alarm and switch functions.
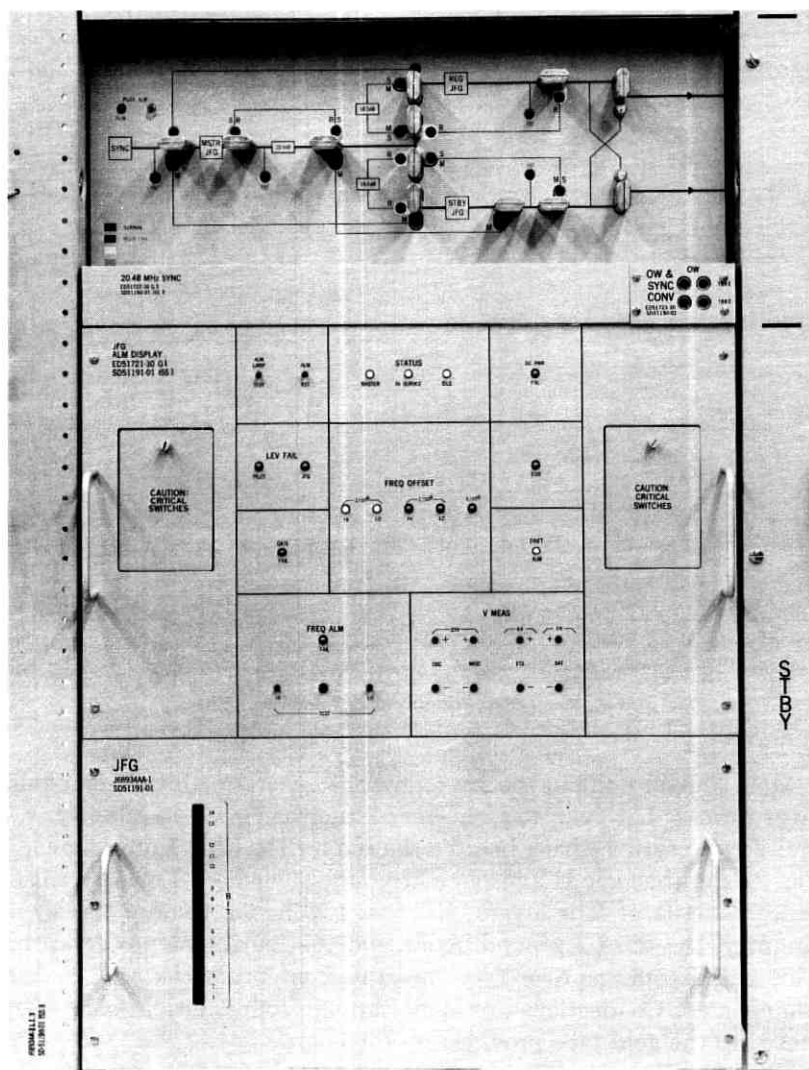
Fig. 23—Jumbogroup frequency generator and patch panel.
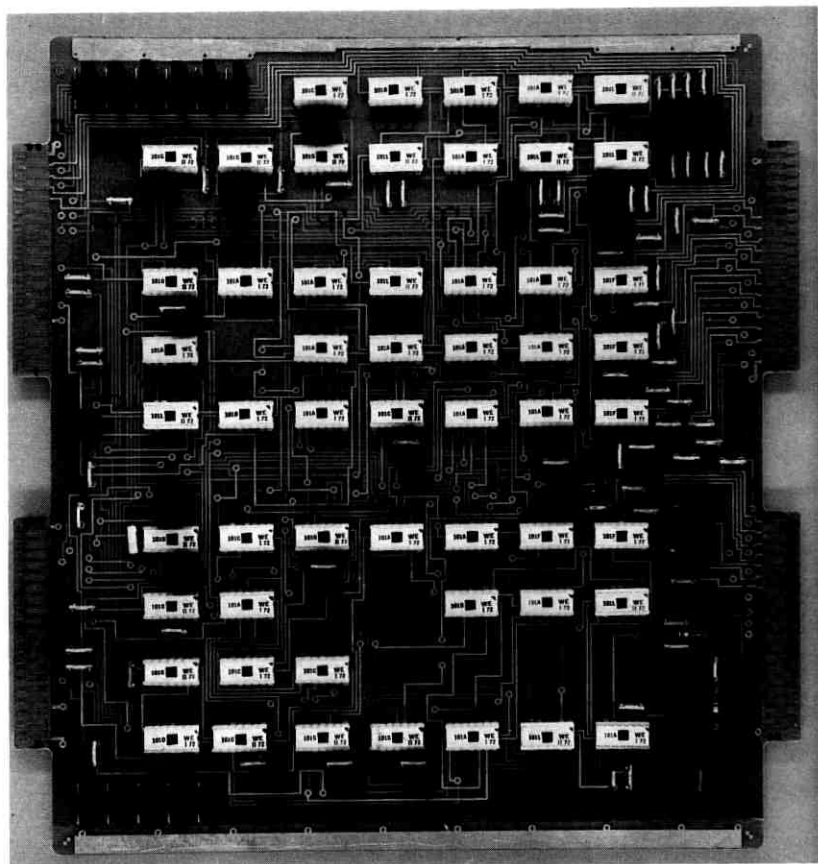
Fig. 24—JFS alarm and switch control logic.

Most plug-in units in the JFS follow the general main-station equipment format; however, to meet very stringent circuit requirements, a multilayer printed wiring board was used for the logic I unit shown in Fig. 24 and the logic II and JFG alarm units. The logic I printed wiring board consists of four layers; the first two layers contain the signal circuitry, the third a ground plane, and the fourth the power wiring. This board contains over fifty integrated circuit packs and discrete components. Connections are made through four multipin connectors mated to the gold tabs provided on the board.

The complete power converter panels are provided, one per JFG. Each converter receives power from a separate office feeder lead. This

ensures that, even if a feeder is accidently damaged or a main fuse operates, the JFS will continue in service.

3.3 Surveillance equipment

The L5 surveillance[9] physical designs cover the widest cross section of types of equipment, ranging from the pure functional types located in manholes to the sophisticated equipment located in the center of activity in a control terminal or terminal-main station. In addition to being influenced by functional design considerations, the latter type of equipment was strongly influenced by human-engineering design considerations and, since it is constantly in the forefront and under scrutiny, was designed to have an aesthetically pleasing appearance. While the repeaters may be the heart of the L5 system, the transmission surveillance system (TSS) equipment is its nervous system, feeding back information on the health and well-being of the system.

3.3.1 Fault-location oscillator and logic gate

The fault-location oscillator (FLO) unit and the 31A gate are the manhole-mounted units of the TSS. A FLO unit is installed in every apparatus case, a unit with eight outputs located centrally in a basic repeater apparatus case (see Fig. 2, Section 1.1) and a unit with four outputs for use with regulating and equalizing repeaters. The FLO is a rugged unit made up of an inner chassis and an outer, fabricated, shell-like housing. The inner chassis supports the face plate and contains the four tone generators (two high frequency, two low frequency), the combining circuitry, and the coaxial switches and associated logic circuitry to turn on the unit. Connections within the unit are made with miniature coaxial connectors and connections for the tone outputs are by way of large Bell System coaxial jacks located in the face plate. The coaxial patch cords used to connect the FLO tone outputs to the input and output of every repeater (through the twin coaxial-jack assemblies located on the apparatus case chassis) are shipped with the apparatus case and provide continuity through the apparatus case for cable acceptance and corona testing. For this reason, the cords are required to be high-voltage designs, even though they are primarily intended to be used at low voltage. The protective, outer covering of the FLO unit consists of a long, thin-walled, rectangular, aluminum housing closed at one end, where it is welded to a rugged, cast, mounting base. This base holds the multicontact connector for input power and logic connections, and contains the large retainer guides that mate with the retainer pins on the apparatus case chassis.

The 31A gate, the logic unit for the FLO's, is used on a one-per-manhole basis and is located in the cross-connect apparatus case (Section 1.1). It consists of a large epoxy-glass PWB assembly housed in a sheet-metal, fabricated can mounted on a hinged shelf that swings out to allow access to wire terminals and protector blocks inside the cross-connect apparatus case.

3.3.2 Line access bay

Rectifiers used to power the FLO units and control circuitry for the FLO units are located in the main station in the line access bay that serves as the interface between the line and the surveillance[9] equipment. Connections to the surveyed functions in the line bays are made with solid-dielectric, coaxial-cable runs to the line access bay where provisions are made to store the excess cable created by the requirement for equal-length, balanced connections. Switched access to the line bays is provided by 1-by-12 dry-reed coaxial switches mounted on a panel together with installer-wired arrays of coaxial jacks located on jack strips designed to be removable from the panel for easy access. "Hairpin-type" plugs are used for interconnections, and the entire panel is recessed behind the bay uprights to protect the plugs from inadvertent disconnects. The recessing is evident in Fig. 17, where the line access bay appears as the third bay in the lineup. The lower part of the bay contains the order-wire-access panel followed by several fixed shelves containing cable equalizers. The last element in the bay is a removable drawer-type storage shelf for the equalizer adjustment unit (EAU) described below.

3.3.3 Surveillance and distribution

The transmission surveillance system is made up of the transmission surveillance center (TSC) and distribution bay combination for control offices along an L5 backbone route and the transmission surveillance auxiliary (TSA) for other than control main stations.

The TSC is a large desk-type console that houses, in its sides, a commercial computer, a transmission measuring test set, and associated equipment. The central portion of the console contains the manual control panel that displays an array of keys and status indicators within convenient reach of an operator. The TSA is virtually a TSC without the computer and is a standard 26-inch wide unequal flange duct-type bay.

Conduction and natural convection were inadequate to transfer heat from the thermal sources within the TSC. To avoid local hot spots,

forced convection is required, and a fan is installed as part of the TSC equipment.

3.4 Peripheral equipment

Associated with a project of the magnitude of an L5 system is a whole series of peripheral items of equipment too numerous to cover in full detail here but of obvious importance to the system and its workings.

Order-wire bays have been coded and, for the first time, make convenient mounting arrangements available to the field where, heretofore, only miscellaneously mountable equipment was provided. In addition, connectorized versions of these bays have been designed to avoid difficulties encountered during the L5 initial installation to accomplish the myriad wiring connections for complete order-wire fan-out in an office.

Multi-mastergroup restoration and zero-loss trunk bay designs involve the use of inexpensive "pseudo-plug-in" construction where simple modules are arranged on a shelf and connected at the rear of the bay with jack-terminated cable arms that are part of the bay wire harness.

Because of the relative inaccessibility of repeaters in manholes and the expense involved in entering a manhole, repeaters that are fully tested before leaving the factory are again tested at field maintenance centers before being dispatched to the manholes. For this purpose, a special repeater test stand and associated warm-up rack were designed for repeater acceptance testing. The test stand simulates the apparatus case chassis details, which make provisions for mounting the L5 manhole plug-in units for verification prior to installation. The verification procedure is speeded up by pre-heating the repeaters on the warm-up rack, which is comprised of five mounting panels, each of which can hold two basic repeaters or one regulating repeater, yielding a total capacity of ten basic or five regulating repeaters. The same retainer guide pins that are part of the apparatus case chassis are used to secure the repeaters on the warm-up rack panels. Powering is by a commercial power supply, and the entire unit is mounted on casters for additional flexibility in those offices designated as maintenance centers.

Required settings on manually adjustable E1 and E2 equalizers are made using the equalizer adjustment unit (EAU) designed as a compact, lightweight, portable test set (see Fig. 25). The unit uses PWB plug-in modules, is designed for ease of assembly, offers accessibility for main-

Fig. 25—Equalizer adjustment unit.

tenance, and is stored in the pull-out drawer at the bottom of the line access bay.

IV. RELIABILITY

With the Bell System providing the major long-haul communications networks for the country, Bell Laboratories designs have always been motivated toward reliability. Systems like the J- and K-carrier systems designed in the 1930's and installed in the 1940's are still providing reliable service. These systems, however, were limited to transmitting 12 two-way telephone conversations over a cable pair, whereas the new L5 carrier system transmits 10,800 conversations per coaxial pair and 108,000 conversations on a fully loaded 22-tube coaxial cable. To accomplish this, new more sophisticated technology was required and to complement this effort a more definitive reliability program was initiated.

We define reliability as ensuring that a given component, equipment, or system will perform a required function, under stated conditions for a needed period of time.

This comprehensive reliability program was instituted almost from the start of the L5 carrier system development and has been proceeding concurrently with it. This overall program covered the following nine functional steps, some of which must still be completed and, therefore, will not be part of this report:

(*i*) Derive early estimates of equipment and system failure rates, mean time to failure (MTTF), and availability.

(ii) Monitor the actual laboratory prototype experience.

(iii) Analyze field-trial production experience.

(iv) Acquire and analyze actual field-trial data.

(v) Review and revise, if required, early estimates based on laboratory and field-trial experience.

(vi) Monitor the Western Electric manufacturing and installation initial route experience.

(vii) Conduct a full-scale reliability study of the initial service route.

(viii) On a continuing basis, obtain and analyze data on all L5 units returned for repair.

(ix) Conduct a reliability study on a subsequent route, installed about two years after the initial route.

The purpose and objectives of this program are to obtain accurate data on the reliability performance of the L5 carrier system as early as possible, to aid in formulating the early system design concepts, to uncover potential problem areas so that corrective action can be expedited before the first system is turned up for service, to compare the actual system reliability performance with the early estimates, and to review the reasons for any major differences, so that more accurate estimates can be made on future systems.

Very early in the development program, a reliability study of the L5 system design was made. As a first step, "black box" failure rate estimates were made on all the repeaters and equipment units. While it is impossible to predict for any individual electronics part either its life or its rate of degradation under known stress, it is possible to treat large populations of such parts on a probabilistic basis with acceptable results. These results can then be related to the statistical behavior of the system. The "black box" failure rates were calculated by summing the failure rates of the components and parts. The component rates used were based on experience in well-defined systems, operating under normal Bell System conditions and environment. For new devices or technology with little or no previous experience, estimated failure rates were derived after consultation with the device or component designers. In these cases, the rates reflected a very conservative estimate.

Using the basic repeater which is the most numerous and least sophisticated of the line repeaters as an example, it was estimated that the failure rate would be about 900 FIT's, with a FIT defined as one failure expected or experienced in 10^9 hours of service. This results in an estimated MTTF of over 125 years for each basic repeater. The

regulating, equalizing, and main-station repeaters being substantially more complex have shorter MTTF estimates.

To obtain overall estimates, two basic system models were selected. Both were 4000 miles long and employed automatic protection switching for the coaxial tubes. In one case, the calculations were based on switching section lengths at a maximum of 150 miles and, for the second, a nominal 120-mile spacing was used, based on the L4 coaxial-system route layout experience. There were some obvious trade-offs in the two models selected. For instance, using the maximum length section resulted in more line repeaters per switching section; however, in a 4000-mile route the number of main stations are reduced from 33 in a route with normal switching section lengths to 27 for the maximum length sections.

An important consideration in the system availability estimates is the average time to repair or restore service. Since it was difficult to arrive at a time that was acceptable, calculations were made based on optimistic and pessimistic restoral times. The optimistic times were defined as an average of 15 minutes for main-station equipment and three hours for manhole equipment, and the pessimistic times as three hours for main-station equipment and six hours for manhole equipment. Recent data on the service restorals for existing coaxial systems show that the average repair times actually being experienced are very close to the optimistic figures.

It is estimated that, for a pair of coaxial tubes in a fully equipped 22-tube cable, with a 1-for-10 automatically switched redundancy, the availability of the electronics and switching contained in a 4000-mile route using the optimistic restoral times would be 99.986 percent for an average outage time of about 1 hour and 15 minutes per year. If the pessimistic repair times are assumed for the same system, the availability would be 99.948 percent per year for an average of about 4.5 hours per year outage.

One very early system question was whether, with double the number of repeaters in a L5 switching section as compared with the L4 system, the ratio of one spare coaxial automatically protecting ten working coaxials was adequate to ensure reliable service. Availability calculations were made on the system model with 120-mile switching sections and optimistic repair times employing 1-for-10 and 2-for-9 protection ratios. With the 2-for-9 arrangement, the probability of three simultaneous failures in a switching section was remote; however, the estimated outage attributable to the more sophisticated switching system resulted in outage times for both strategies being

relatively close. Based on previous experience, cable failures constitute a greater outage hazard in terms of system unavailability than do the electronics. Therefore, using additional coaxials within the same cables as protection for most cable faults would be counter-productive. Taking into account these factors and the revenue lost by dedicating a second coaxial for protection, the 1-for-10 protection ratio was continued for the L5 system.

The field trial of the L5 carrier system started on July 1, 1970 and continued through June 1972 on a 125-mile cable route between Cedarbrook and Netcong, N. J., shown in Fig. 26. The purpose of the field trial was to test the L5 as a complete system operating in the field environment before going into full-scale production and regular service. The 240 basic repeaters and 48 regulating repeaters required for the trial were constructed by Western Electric on a special basis because the quantities exceeded Bell Laboratories model shop capabilities. Since problems in the field can often be anticipated from experience gained during manufacture, a data acquisition program was initiated by Western Electric for the field trial equipment and, eventually, standard production.

A reliability study was made on the performance of the field trial route. This was intended as prelude to a study of the initial L5 service route between Lillyville, Pa. and Hillsboro, Mo. Data were collected on all failures of equipment and all system outages. A thorough investigation was made on any units that failed. A case in point was the failure of a basic repeater traced to the earth-ground filter. After further analysis, including the X-ray photograph shown in Fig. 27, it was found that high-voltage shorting was caused by a capacitor unit assembled off-center without outer insulating material. To prevent this in the future, high-voltage test screening was specified and tighter quality controls were instituted by the supplier. Failure rate calculations were made based on the field data; however, it was recognized that the population was small, the equipment was not from standard production, and, because of the level of field activity, it was difficult to keep accurate data on every single occurrence.

Our experience with previous systems such as the L4 coaxial system and the L5 field trial pointed out the need for even greater emphasis on obtaining accurate data on the manufacturing testing results, the initial defect rates in the field, the infant mortality rates (the infant mortality period is defined as the first six months of service), and, finally, the steady-state system reliability performance more commonly known as the normal use period of the "bathtub" curve. To accomplish
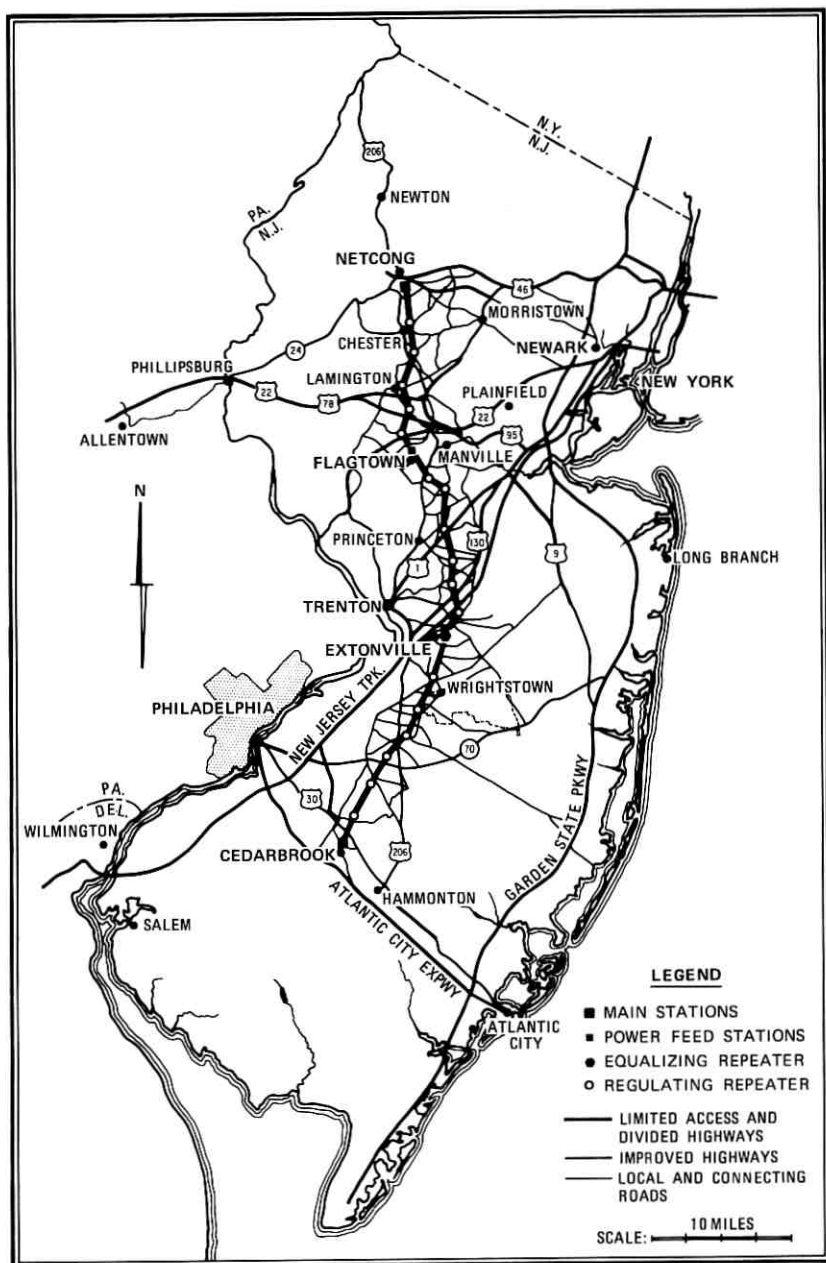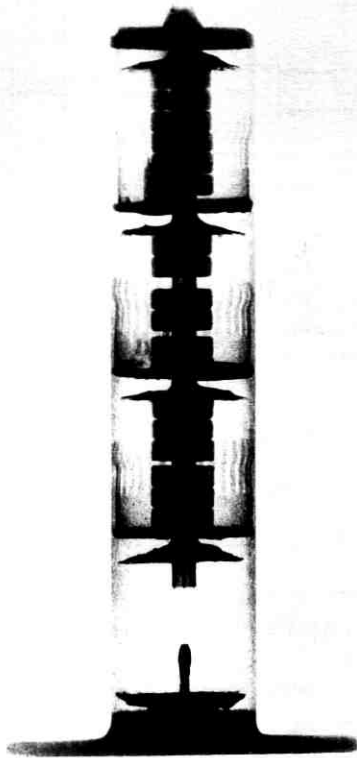
Fig. 26—L5 field trial route.

Fig. 27—X-ray of earth-ground filter.

this, an elaborate data collection program was initiated. Subunits such as coded amplifiers and even HIC substrates were serial-numbered. Special envelopes and data forms were attached to each unit at the start of manufacture. As the unit was processed through the various stages of manufacture, the test results were recorded on punched tapes and stored in the envelopes. If components are removed at some stage, they are stored in the envelopes and the data recorded on the forms. When, for instance, a repeater is ready to ship, a complete pedigree has been established for analysis and future reference. The components removed are forwarded to the experts for analysis to determine the failure mechanism. Through this program, we are able to detect failure trends and initiate corrective action.

An important part of the reliability program is to establish the initial field defect rate. In Fig. 28, a block diagram traces the pro-
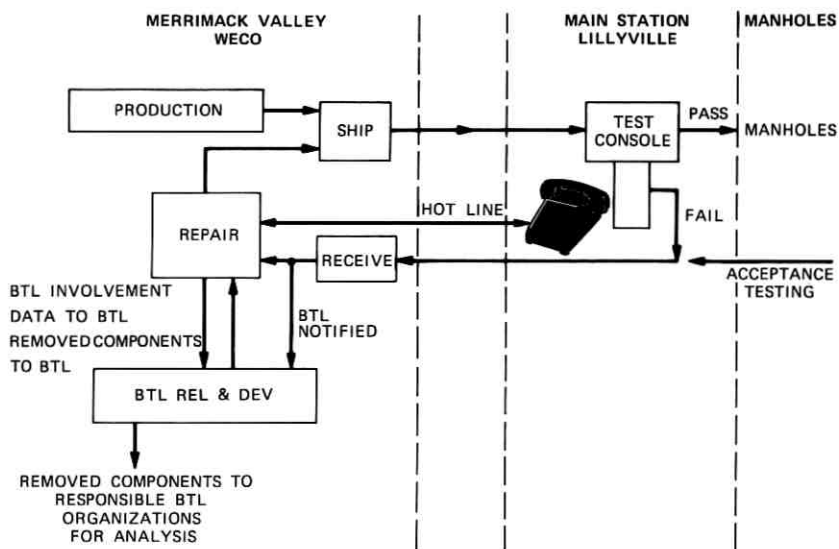
Fig. 28—Initial L5 route returns repair-reliability program.

cedures set up to process units returned for repair. Note that a "hot line" was established by Western Electric to expedite the replacement of any units returned for repair so that the service schedules would be met. When units are returned for repair, the designers are able to participate in diagnosing the problems. Data are recorded on all repairs, and the components removed are kept for further analysis. For the initial route, about 3200 basic repeaters were tested and installed in manholes. Approximately 1.5 percent were returned for all reasons. After further testing, it was determined that about 35 percent of these met all test requirements, resulting in a 1-percent initial return rate.

On January 3, 1974, the L5 route between Lillyville and Morgantown, Pa., was put into regular service, and the L5 field reliability program was started. On January 25, the section from Morgantown to Hillsboro, Mo., was put into service, and this section was added to the reliability study. Before the start of service, visits were made to all offices on the route to discuss the objectives of the study and the type of data required. During the early service period, each office is contacted on a weekly basis to report all field problems, even those not affecting reliability. The purpose is to get immediate feedback on the system performance. A complete listing of all equipment in the initial route by serial number was made so that the hours of service necessary for availability calculation can be accurately computed.

L5 COAXIAL SYSTEM RELIABILITY STUDY
LILLYVILLE - HILLSBORO
OUTAGE AND EQUIPMENT FAILURE

STATION REPORTING **Lillyville, Pa**
DATE **11-2-73**

I. **SWITCH TO PROTECTION COAX** (ONE MINUTE OR MORE)

☑ AUTOMATIC ☐ MANUAL AUTO LINE TERMINATE ☐ YES ☐ NO

LINE NO REGULAR **L503** PROTECTION **L501**

DATE·TIME ON **11-2** _____ **1620** DATE/TIME OFF **11-2** _____ **1835**
DATE · TIME DATE · TIME

REASON LINE FAILURE ☑ MAINTENANCE ☐ RESTORATION ☐
(EXPLAIN IN II) (EXPLAIN IN III)

II. CAUSE OF LINE FAILURE

☑ LINE EQUIPMENT FAILURE
☐ CABLE FAILURE
☐ TERMINAL EQUIPMENT
☐ H V CONVERTER FAILURE
☐ OTHER CAUSES
SPECIFY _____

III. MAINTENANCE

☐ MODIFICATION
☐ OTHER

IV. OUTAGE TIME CAUSED BY
(IN ADDITION TO SECT II)

☐ PROT COAX IN USE
☐ PROT COAX IN USE FOR RESTORATION
☐ SWITCHING FAILURE
☐ TANDEM FAILURE
LOCATION _____
(IF KNOWN)
☐ SIDE LEG FAILURE
LOCATION _____
(IF KNOWN)

V LINE AND TERMINAL EQUIPMENT FAILURE *

CODE AND LIST **J68919AA-1 Li;db3** L NO **1375** LOCATION **2081-0270**
(MANHOLE DESIG. OR MAIN STATION)

DATE IN SERVICE **1-8-73**

DATE/TIME OF FAILURE **11-2** _____ **1620** * SEE INSTRUCTIONS FOR
DATE TIME EQUIPMENT TO BE REPORTED
DATE/TIME RESTORED **11-2** _____ **1835**
DATE TIME
FAILURE SERVICE AFFECTING? ☐ YES ☑ NO

NATURE OF FAILURE CODE _____ OTHER _____
(SEE CODE ON OTHER SIDE)
ENG COMPLAINT # _____

VI LINE OUTAGE

LINE(S) FAILED L# _____ L# _____ L# _____ DATE/TIME _____
DATE TIME

LINE(S) RETURNED TO NORMAL DATE TIME

L# _____ DATE/TIME _____ _____
L# _____ DATE/TIME _____ _____
L# _____ DATE/TIME _____ _____

MAIL TO MR W C WESTPHAL, ROOM 3B13
BELL TELEPHONE LABORATORIES
1600 OSGOOD STREET
NO ANDOVER, MASS 01845

REMARKS

Fig. 29—Return data form.

A data form shown in Fig. 29 is filled out and returned to Bell Laboratories for failures of any type, even if service was not affected, and for the use of the spare line facilities for protection and for maintenance. These forms are returned to Bell Laboratories weekly for compilation. Arrangements have been made so that complete data will be recorded by Western Electric when the units are returned for repair. At the time this paper was written, the system was in service

for a very short period of time and the field data were limited; therefore, no reportable results were available. This reliability study is intended to continue for a minimum of one year.

As part of our continuing transmission system reliability program, the general performance of the L5 system will be continually monitored by our computer-aided reliability program (CARP), which utilizes data received from the service centers on all units returned from any route for repair. In addition, in about two years we intend to conduct a reliability study similar to the one presently in operation on another route.

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L5 SYSTEM:

Ultralinear Transistors

By F. A. D'ALTROY, R. M. JACOBS, J. M. NACCI,
and E. J. PANNER

A family of ultralinear npn transistors has been developed for use in the L5 coaxial-carrier system. These 3-GHz devices are characterized by extremely low distortion and noise figure. The transistor comprises an interdigitated base-emitter structure with a heavily doped base grid connected to the peripheral base metal contact. The emitter contact is overlaid on the base-emitter region. Contact metallurgy consists of a platinum silicide, titanium, platinum, and gold system. The transistor is a highly reliable device and meets all the performance requirements of the L5 system.

I. INTRODUCTION

To realize the circuit performance anticipated by raising the coaxial carrier's highest message frequency from the 17.5 MHz of L4 to the 60.5 MHz of L5, it was necessary to develop a new family of npn transistors, the 76 and 77 types. The gain bandwidth product f_T had to be nominally 3,000 MHz; three-tone third-harmonic distortion M_{3E} had to be less than -88 dB below 1 mW and the high-frequency noise figure (NF) kept below 2.4 dB at 30 mA. Power dissipation was to be 1.7 watts, which amounted to a considerable improvement in the state of the art.

The 76-type transistors are used in the equalizing and regulating repeaters and the 77-type in the basic repeater. The 77's are used as matched pairs to lower intermodulation distortion and the noise figure.

One basic design is used to meet all of the equalizer and repeater requirements. Six codes were developed in all, each of which was chosen for a specific purpose, such as low noise for the input stage of the basic repeater, and low distortion for the output stage. Typical characteristics for two codes are given in Table I.

Table I

	77E Typical	77D Typical	
$V_{(BR)CEO}$	20	23	V
h_{FE}	100	90	
f_T	2200	3000	MHz
NF	2.2	2.7	dB
C_{ob}	3.6	3.6	pF
M_{3E}		-90	dB

II. STRUCTURE

Some early studies indicated that of the common geometries of power transistors, an interdigitated structure would have the best distortion performance; thus, the present devices are made with interdigitated base and emitter diffusions, but with an overlaid emitter contact. Base current is carried by the heavily doped base grid, under the overlaid contact, and out to metallized regions at the sides. See Fig. 1. This structure avoids metallizing problems caused by tight metal-contact tolerances, and it minimizes the total area. Emitter bonds are made over the active region, eliminating the emitter pad and its parasitic capacitance. The emitters are 2.5 μm wide and 105 μm long.

The epitaxial thickness is a compromise between being large enough to provide sufficient series collector resistance to minimize second-breakdown problems and yet not so large as to degrade device performance.

III. ENCAPSULATION

Following the practice set by the L4[1] system, the transistors are encapsulated in a small metal-ceramic package especially designed for rf use. It comprises a kovar-beryllia structure that minimizes parasitic capacitance and lead inductance and provides a relatively low thermal impedance of 30°C/watt. The latter characteristic is important if high reliability and low distortion (see Fig. 2) are to be obtained, since the output transistors in the basic repeater will operate at 1.7 watts in a peak ambient of 85°C.

IV. DISTORTION

The principal contributor to system noise from the output stage of an amplifier is third harmonic distortion, M_{3E}. In a transistor, this noise arises from two main sources: from the essential nonlinearity of
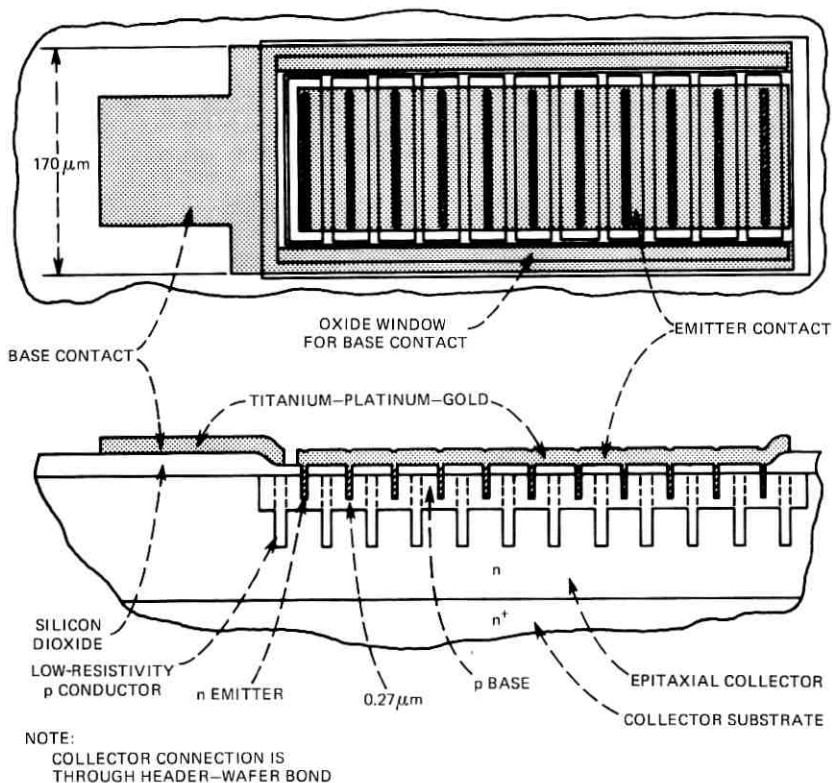
Fig. 1—Structure of L5 transistors.

the emitter current-voltage characteristic and from the falloff of f_T with current[2] owing to conductivity modulation of the collector-junction space-charge region, or Kirk effect.[3] The emitter nonlinearity is the same exponential behavior of the forward-biased diode used, for example, in mixers and for harmonic generation. The nonlinearity is reduced by increasing the total emitter current. Poon[2] has shown that if the emitter area and collector current (effectively the emitter current) are both doubled, M_{3E} decreases by 12 dB (see Fig. 3). If the current density becomes too large, however, distortion arising in the collector starts to degrade the device performance. Whenever the current density exceeds qvN, where q is the electric charge, v the carrier velocity, and N the density of fixed impurities at the collector, the collector-base junction will move in towards the collector substrate and cause f_T to decrease. According to Poon,[2] in the L5 case there is a
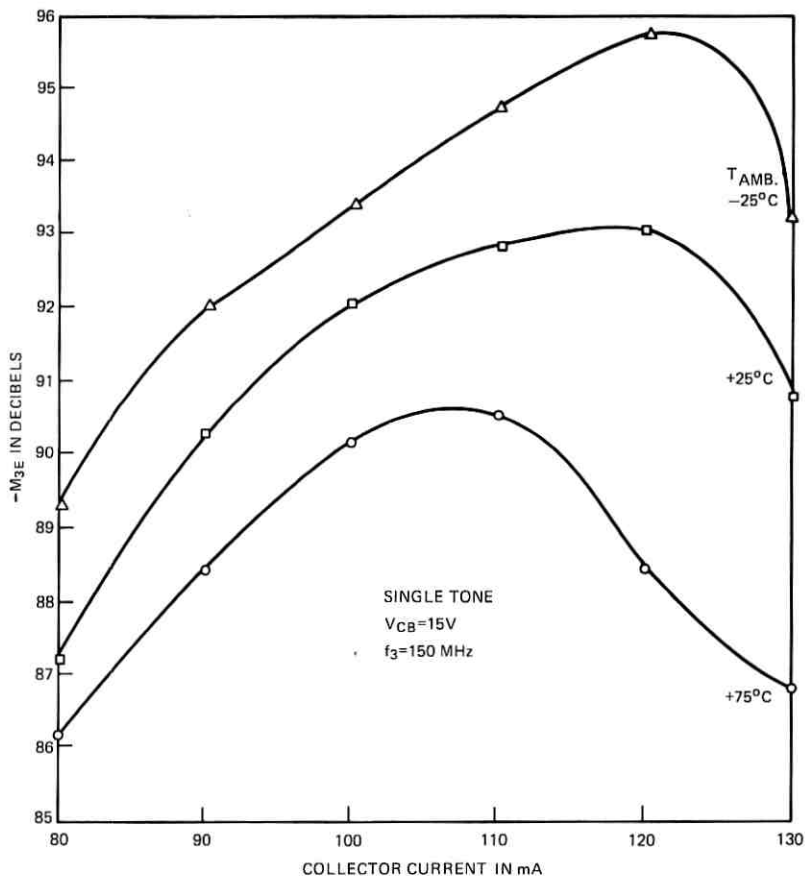
Fig. 2—Temperature dependence of M_{3E} for L5 transistors.

limiting condition given by

$$M_{3E} = -20 \log_{10} \left\{ \frac{R_L 1.2 \times 10^4 \times f_T}{\frac{\partial^2 f_T}{\partial I_c^2}} \right\},$$

where R_L is the load resistor and I_c is the collector current.

It can be seen that any mechanism, such as the Kirk effect, which increases the curvature of the f_T vs I_c characteristic, will contribute to harmonic distortion. This means, of course, that as the emitter current is increased to minimize the emitter-distortion term, the collector area must increase proportionately. At some point it ceases to be economic
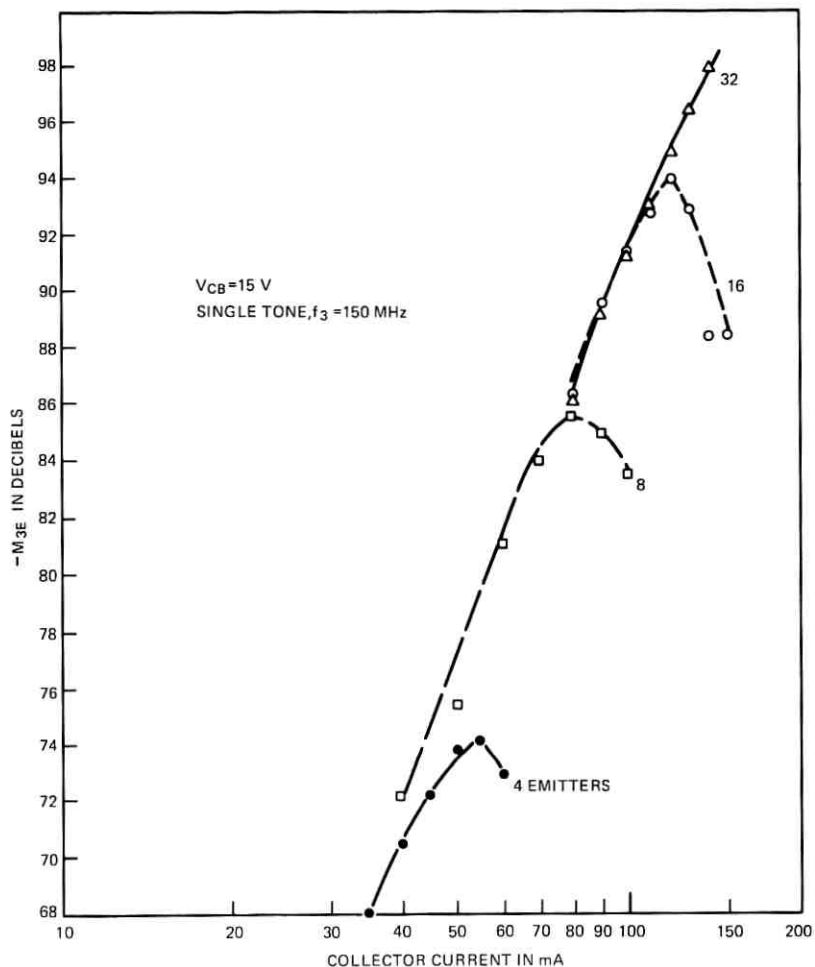
Fig. 3—Harmonic distortion as a function of collector current for several device sizes.

to increase the emitter current because both capacitance and dissipated power become excessive. For the transistor size chosen, a current of 110 mA provides the lowest distortion.

V. SUSTAINING VOLTAGE

The sustaining voltage for the L5 transistor turns out to be bounded at both limits. At the lower limit, it is necessary to keep above the supply voltage, so a value of 18 volts minimum is set. At the upper

limit the doping may become too low and a Kirk effect may occur if the value of the sustaining voltage is made too high. This result can be seen from the following relationships[4]

$$V_{BD} = 1.71 \times 10^9 \left(\frac{N_{BC}}{x_j} \ln \frac{N_o}{N_{BC}} \right)^{-0.364}$$

and

$$V_{CEO} = \frac{V_{BD}}{(\beta)^{1/n}},$$

where

N_{BC} is the collector epitaxial doping
N_o is the surface impurity concentration
x_j is the junction depth
n is a constant $\simeq 4$ for the transistor concerned.

A practical upper limit is 28 volts.

VI. GAIN BANDWIDTH PRODUCT

One of the basic differences between L4 and L5 transistors is brought about by the increase of the highest message frequency from 17.5 MHz to 60.5 MHz. To accommodate the increased frequency, it is necessary to design the transistor to have a nominal gain bandwidth cutoff, f_T, of 3 GHz. The solution to f_T is given by

$$\frac{1}{2\pi f_T} = \tau_e + \tau_b + \tau_x + \tau_c,$$

where

τ_e = emitter time constant
τ_b = base time constant
τ_x = collector depletion region time constant
τ_c = collector capacitance time constant.

In the range of large collector-current densities, i.e., $J_c > J_2$,

$$\tau_e = \frac{a_o KT}{qJ_c} C_{te}$$

$$\tau_b \approx \frac{(W_b + W_c)^2}{nD_n} \left\{ 1 - \left[\frac{W_c}{W_b + W_c} \frac{J_2}{J_c} \right]^2 \right\}$$

$$+ \frac{W_b + W_c}{v} \left\{ \frac{J_2}{J_c} \right\} \left\{ 1 - \frac{W_c}{W_b + W_c} \right\} \frac{J_2}{J_c}$$

$$\tau_x \approx 0$$
$$\tau_c \approx 0,$$

where

K = Boltzmann's constant
T = absolute temperature
q = electronic charge
J_c = collector-current density
C_{te} = emitter capacity
W_b = base width
W_c = collector width
D_n = diffusion constant for electrons
$J_2 = \dfrac{-V_{cb} + V_o}{\rho_c W_c}$
V_{cb} = applied voltage
V_o = built-in junction potential
v = carrier velocity

Since τ_b dominates the frequency response, the necessity of keeping the current density low can be readily seen.

Under this condition, τ_b, τ_x and τ_c reduce to

$$\tau_b = \frac{W_b^2}{nD_n} + \frac{W_b}{v_s}$$

$$\tau_x = \frac{X_{mo}}{2v_s} \left\{ \frac{1 - J_c/J_2}{1 - J_c/qv_sN_{BC}} \right\}^{\frac{1}{2}}$$

$$\tau_c = \epsilon\rho_c \left\{ \frac{W_c}{X_{mo}} \frac{1 - J_c/qv_sN_{BC}}{1 - J_c/J_2} \right\}^{\frac{1}{2}},$$

where

v_s = scattering limited velocity
X_{mo} = depletion layer width
ϵ = dielectric constant.

Once the current density has been chosen to be suitably low, the only available design factor for lowering f_T is the actual base width. To achieve an f_T of 3 GHz, it is necessary to lower the base width to 0.27 μm nominal.

VII. NOISE FIGURE

Neilsen's relation[5] for the noise figure, NF, provides a simple means of determining the choice of design parameters for minimizing the noise figure.

$$\text{NF} = 1 + \frac{r_b'}{R_g} + \frac{r_e}{2R_g} + \frac{(1 - \alpha_o)\left[1 + \left(\dfrac{f}{\sqrt{1 - \alpha_o f_\alpha}}\right)^2\right](R_g + r_b' + r_e)^2}{2\alpha_o r_e R_g},$$

where

R_g = generator impedance of 50 ohms

r_e = emitter resistance = KT/qI_e

f_α = frequency at which α is 0.707 of its dc value

I_e = emitter current.

It is apparent that once current, gain, and operating frequency have been chosen, r_b' is the only parameter at our disposal and, in fact, for the L5 transistor, the r_b' term is the largest contributor to the noise figure. The term r_b' is made up of two parts. One is the lateral resistance under the emitter stripe, which is minimized by using extremely narrow emitters 2.5 μm in width. The outer component consists of the series resistance to the external base contact. This resistance is reduced to as low a level as possible by the use of a very heavily doped base grid.

VIII. RELIABILITY

During development of the 76- and 77-type transistors, several hundred devices were subjected to accelerated aging of both shelf-temperature and power at junction temperatures up to 300 °C. Analysis of the data predicts that the transistors will have a failure rate of considerably less than 10 fits at an operating temperature of 150 °C. Under these conditions, it is no longer reliability that is the determinant for system success, but rather, the temperature dependence of the various device parameters.

IX. CONCLUSION

A highly reliable ultralinear family of transistors has been developed for coaxial-carrier-system use. Its high gain-bandwidth product, low noise, and low third-harmonic distortion allow the L5 system to successfully meet its operational requirements with an upper master-group frequency of 60.5 MHz.

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L5 SYSTEM:

Transmission Networks and Magnetic Components

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The development of transmission networks and magnetic components for the L5 system represents the largest network development project of its type ever undertaken within the Bell System. Over 200 different coded designs of networks, requiring in excess of 40 man-years of effort, were required to meet the frequency-selective and signal-shaping requirements of the system. Despite this effort, neither systems requirements nor systems schedules could have been met without significant contributions from allied technologies. This article identifies those technologies and describes design techniques that have advanced the state-of-the-art capabilities in transmission network and magnetic component design.

I. TRANSMISSION NETWORKS

1.1 Introduction

As in other analog systems, transmission networks of the L5 Coaxial-Carrier Transmission System perform the indispensable functions of frequency selection and signal shaping. Without the filtering functions provided by certain of these networks, the basic multiplexing arrangements on which the system depends could not be realized, nor could the various fault-locating, equalizing, regulating, and switching pilots be effectively separated from the message portion of the line signal. Similarly, without use of the wide variety of equalizers provided, the amplitude distortions introduced as the signal traverses the line could not be as effectively corrected. Examples of such equalizers include *line-build-out* (LBO) *networks* to compensate for cable spans shorter than the nominal one-mile spacing, *fixed equalizers* to compensate for time-invariant and predictable effects, and *adjustable equalizers* to reduce the unpredictable, statistical variations of the system.

To complete a listing of system functions performed by the transmission networks of the L5 system would require a tabulation of all major system functions. Over 200 transmission networks, requiring in excess of 40 man-years of effort and representing the largest network development of its type ever undertaken within the Bell System, were developed for L5. Despite this large and sustained effort, the required designs could not have been produced without significant advances in computer-aided design, precision measurement, magnetic components, and monolithic crystal filters. The purpose of the transmission networks section of this paper is to identify those technologies that have influenced network developments for the L5 system and to describe certain techniques that have advanced the state-of-the-art capabilities in network design.

1.2 Network technology

The majority of networks for the L5 system were developed using classical network synthesis procedures. Insertion-loss synthesis techniques were used exclusively for filter realization, Bode's semi-infinite slope approximation for LBO networks, cable equalizers and artificial lines, and constant-resistance bridged-T networks, in various topologies, for fixed and variable amplitude equalizers. State-of-the-art advances stemmed primarily from the stringent requirements imposed by the system rather than from network theoretic developments. Strong interaction between systems and network engineers, in frequency allocation and the setting of reasonable guard bands, assisted greatly in the problem of filter realizability. In practically all cases, developing techniques for reducing circulating currents, and developing practical design procedures for the compensation of parasitics were a necessary adjunct to the physical realization of manufacturable networks.

1.3 JMX modulation filters

The development of modulation filters for the jumbogroup multiplex terminal (JMX)[1] is typical of that for the many filter designs produced for the L5 system. Figure 1 is a simplified block diagram of the JMX terminal as viewed by a networks designer. The modulation scheme is typical of most analog multiplex systems except for the multiple-modulation steps. Although the double and triple steps of modulation increase the number of filter designs required, they reduce the complexity of the individual filters. The final circuit and networks objectives were the result of extensive discussions between systems and
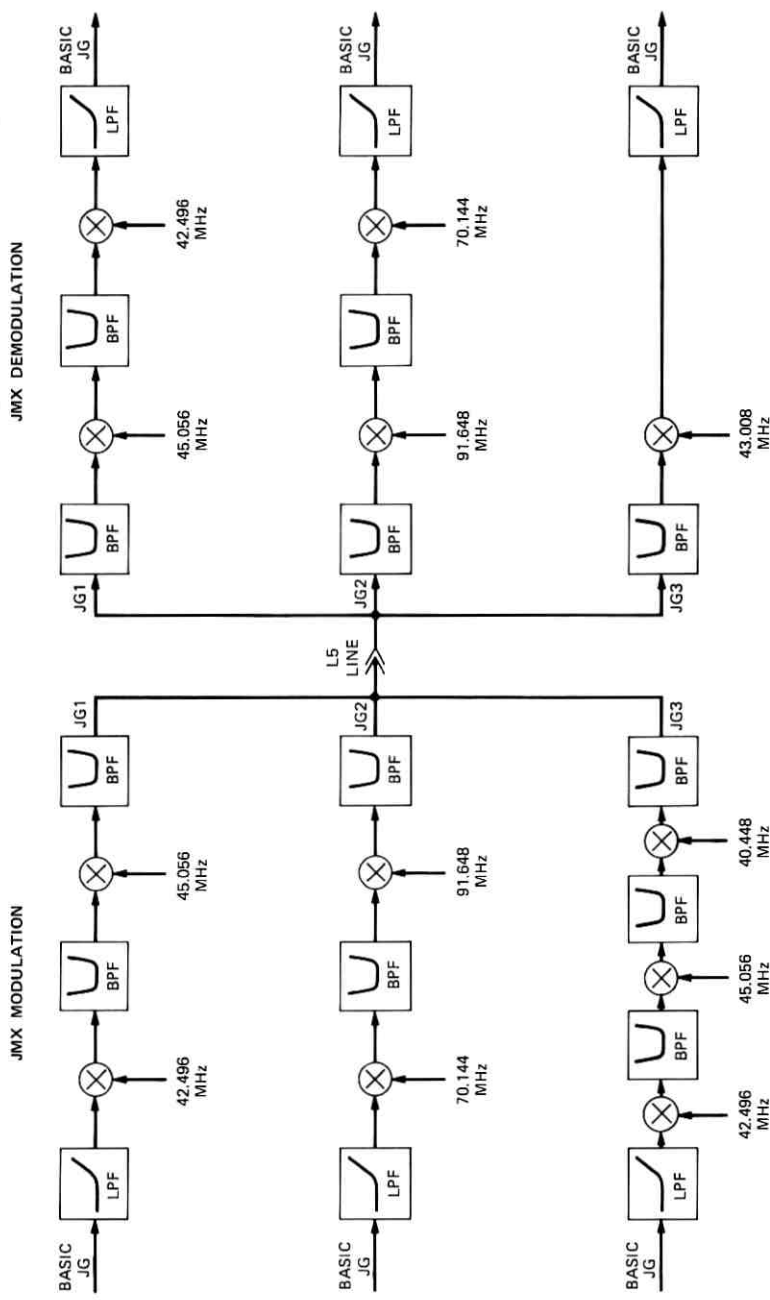
JMX DEMODULATION

JMX MODULATION

Fig. 1—Simplified block diagram of JMX terminal emphasizing transmission networks.

networks engineers with the goal of optimizing the trade-off between frequency-spectrum utilization and the setting of adjacent jumbogroup guard bands.

As a result of such discussions, typical filter transition bands approximated a reasonable 10 percent, which allowed the use of small slug-tuned, air-core inductors having a maximum quality factor (Q) of approximately 150. All 14 of the filter designs—10 complex bandpass, 3 low-pass, and 1 low-pass-high-pass filter—are equal-ripple insertion-loss designs. The degree of the filters ranged from a minimum of $n = 7$, for a simple low-pass filter having a relatively low stopband objective and a wide transition band, to a maximum of $n = 20$, for a more difficult bandpass design having an 80-dB discrimination objective across the unwanted sideband. Figures 2 and 3 show the schematic diagram and measured characteristics, respectively, of one of the more complex bandpass filters.

The initial design phases were concerned with the optimization of the filter designs for practical element values, low passband delay distortion, good out-of-band performance, and acceptable inband return loss. Versatile insertion-loss synthesis, pole-placing, and analysis programs were available.[2] Some of the features of the programs found particularly useful during the design stage were: (i) the ability to choose arbitrary stopbands with Chebychev passband behavior, (ii) the ability to order loss peak removal to optimize element values, and (iii) the automatic computation of network characteristics subsequent to the solution of the synthesis problem.

Extensive use was also made of available programs to minimize the delay distortion contributed by the modulation filters for any future application of the JMX terminal to data systems. Theoretical cutoffs were placed as far into the transition region as possible without unduly increasing the degree of the filters or making the element values impractical. This modification of the designs achieved up to a 50-percent reduction in delay distortion over the desired portion of the passband with only a 10-percent increase in the complexity of the filters.

As a result of the JMX modulation arrangement, the carrier, in many cases, fell close to the passband edges, making it impossible (with the extended theoretical cutoffs) to adequately suppress the carrier leak. Quartz crystals, as shown in the schematic of Fig. 2, were therefore introduced in various shunt branches to suppress the unwanted carrier. Highly stable, high-Q, AT-cut crystals, designed for maximum suppression of unwanted responses, were used. Other components
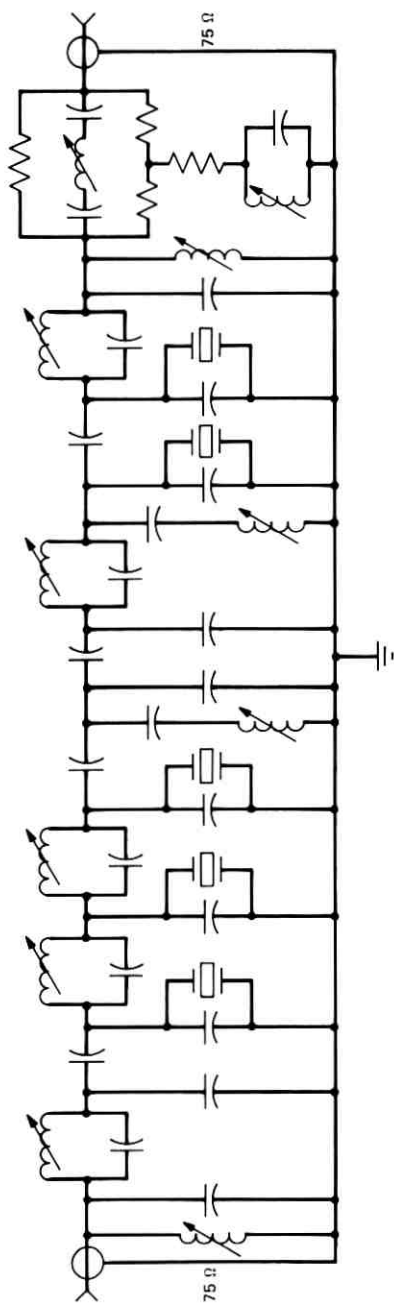
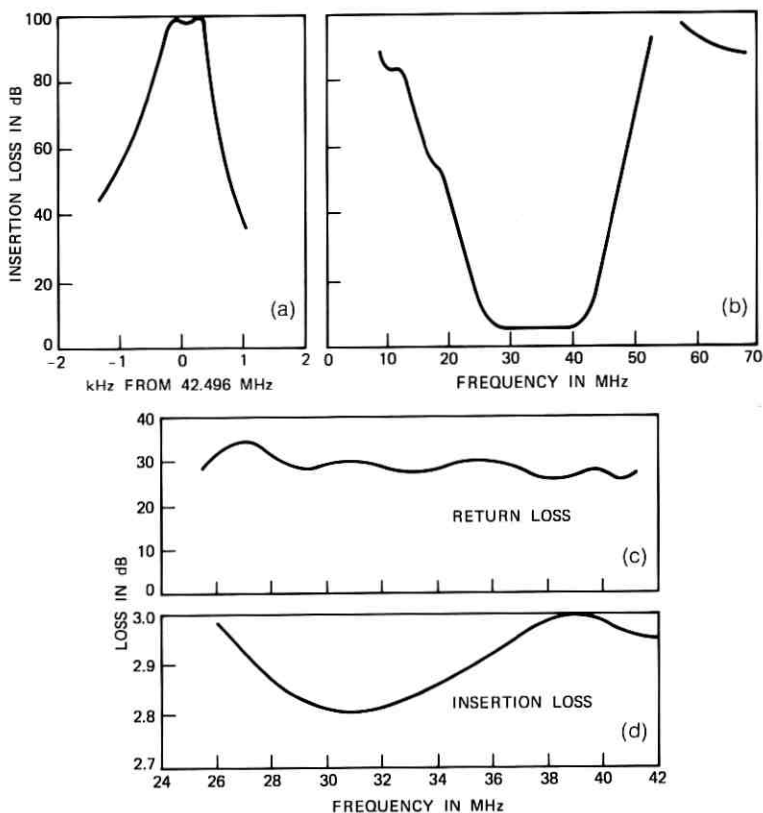Fig. 2—Network schematic of typical JMX bandpass filter.

Fig. 3—Characteristics of typical JMX bandpass filter. (a) Component of loss contributed by crystals at 42.496 MHz. (b) Stopband loss. (c) Return loss. (d) Passband loss.

included adjustable air-core inductors and dipped-mica and tubular capacitors. These components were mounted on printed wiring boards (PWB's) and packaged in sealed enclosures.

Laboratory development of the JMX networks was a difficult and lengthy process because of the wide bandwidths and high frequencies involved. Precision scanning network analyzers were used widely in the initial stages of development, but in all instances the computer-operated transmission-measuring set[3] (COTMS) was used for final characterization.

1.4 Monolithic crystal bandpass filters

Monolithic crystal filters[4] have found wide application in the L5 system as narrow-band pick-off filters and as wide-passband band-

elimination structures. The monolithic crystal filter was first introduced in the Bell System plant as simple two-resonator bandpass filters for carrier-supply applications[5] in the L4 coaxial system. For the L5 system, over 40 designs, ranging in frequency from less than 3 MHz to nearly 100 MHz, were required for both bandpass and band-elimination applications. To cover this frequency range, fundamental and third-overtone crystal designs were necessary, and multiresonator filtering arrangements were provided to meet the high out-of-band suppression requirements.

The monolithic crystal filter, in its most elementary form, consists of a quartz plate with an array of two-electrode pairs placed on its major surfaces, with each pair establishing a resonator system. The resonators are acoustically coupled via the elastic quartz plate. If the quartz plate, electrode pairs, and acoustic coupling are properly proportioned, a self-contained, two-port piezoelectric device, having prescribed filtering characteristics, results. Higher-order filter functions may be obtained by deposition of multielectrode pairs on an extended quartz plate, with adjacent resonators acoustically coupled. As an extension of this arrangement, multiple quartz plates may be used, with adjacent resonators (on facing quartz plates) coupled electrically (capacitively). A third alternative, the composite monolithic crystal filter, uses LC filter sections to couple individual two-resonator monolithics. The latter approach was used exclusively in the design of more than 30 monolithic crystal bandpass filters for L5 applications. Impedance-matching sections necessarily were included in these designs to match the impedances of the coupled-resonator designs to specified source and load impedances. This combination of LC coupling networks, LC impedance-matching networks, and two-resonator monolithics led to the concept of the monolithic crystal filter (MCF) as a two-port, building-block device available for inclusion in some larger frequency-selective network.[6,7]

Before describing those methods of design peculiar to the composite MCF, a more detailed summary of methods of deriving higher-order filter functions is required. Figure 4 shows the three different techniques described above. In each instance, an impedance-matching network is indicated at the input and output ports of the networks. These may or may not be required. Figure 4a shows the multiresonator MCF on a single piece of quartz with the interresonator coupling supplied by the built-in acoustic coupling properties of the device itself. To avoid acoustic coupling of certain unwanted modes, the approach indicated in Fig. 4b is sometimes advisable. In this instance, the single plate of Fig. 4a has been split into two parts and the facing
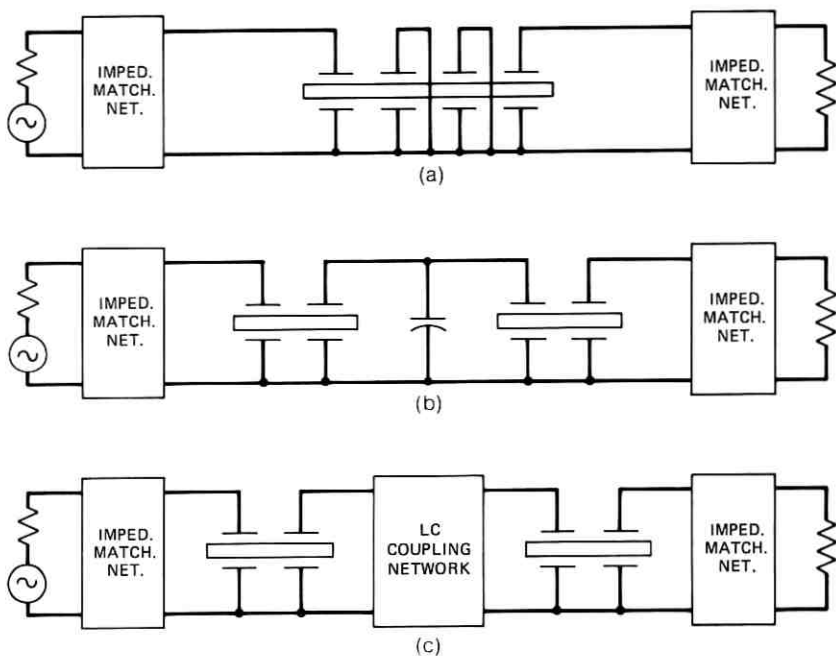
Fig. 4—Methods of deriving higher-order filter functions. (a) Multiorder mono-lithic filter with acoustic interresonator coupling. (b) Multilithic filter with acoustic and electrical interresonator coupling. (c) LC network coupling of two identical MCF's.

pairs of electrodes electrically coupled by the capacitor are shown in the figure. The present design of the A6 channel-bank filters[8] uses this technique. The third method of realization is represented in the schematic of Fig. 4c. This schematic emphasizes the concept of the MCF as a two-port, building-block device employing supplementary LC filter sections (coupling networks) and impedance-matching networks for realization of a particular transfer function. In all three examples shown in Fig. 4, the classical work of Dishal[9] provides closed-form solutions in the design of narrow-band bandpass filters to exact amplitude characteristics.

In view of the importance of the building-block concept to the design of composite MCF's, devices and filter sections required for filter realization are shown in Figs. 5 and 6. Emphasizing the role of the MCF as a two-port device in composite-filter design, schematics for the basic and reversed-phase connections of a symmetrical, two-resonator MCF and their equivalent circuits are shown in Fig. 5. The associated LC networks required in a composite filter design are shown

in Fig. 6. These include both impedance inverters and impedance-matching networks. The impedance inverters are shown in their T- and pi-configurations and consist of positive and negative capacitive elements. In the final network design, the negative capacitances shown in the figure are either absorbed in positive capacitances existing within the network, or are approximated in a narrow-band sense by inductances.

The impedance-matching networks of Fig. 6 are shown in their L-configurations. It is important to note that the shunt elements of these networks must also be designed to "absorb" the input or output static capacitances of the associated MCF's.

Given the equivalent circuits of Figs. 5 and 6, design techniques applicable to multiresonator bandpass filters become readily apparent. A familiarity with the theory of coupled-resonator design, however, is assumed. References 9, 10, and 11 provide excellent introductions to the theory. Starting with a general n-resonator structure, the equivalent, composite MCF is easily derived by a series of network decompositions as shown in Fig. 7. A filter design having equal inductance values throughout the structure is indicated. As shown in the figure, impedance inverters, having element values corresponding to the shunt capacitances of the structure, have been introduced into the circuit at each of the shunt branch points. Certain sections of the
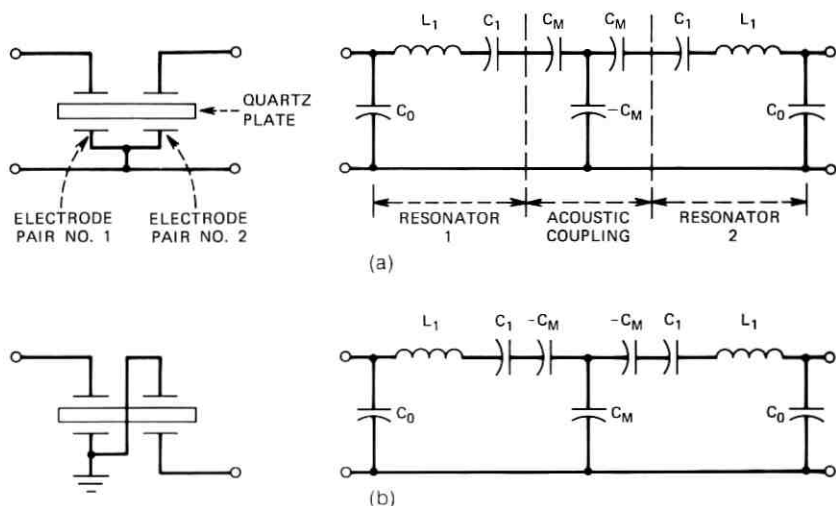

Fig. 5—Schematics and equivalent circuits of monolithic crystal filters. (a) Basic bandpass filter. (b) Reversed-phase bandpass filter.

IMAGE PHASE SHIFT = $-90°$ IMAGE PHASE SHIFT = $90°$

EQUIVALENT NARROW–BAND IMPEDANCE INVERTERS

LOW PASS

$R_2 > R_1$

$$L_1 = \frac{1}{\omega_0} \sqrt{R_1 (R_2 - R_1)}$$

$$C_1 = \frac{1}{\omega_0 R_2} \sqrt{\frac{R_2}{R_1} - 1}$$

HIGH PASS

$R_2 > R_1$

$$L_1 = \frac{R_2}{\omega_0} \sqrt{\frac{R_1}{R_2 - R_1}}$$

$$C_1 = \frac{1}{\omega_0 \sqrt{R_1(R_2 - R_1)}}$$

BANDPASS

$R_2 > R_1$

$$L_1 = \frac{R_2 \cdot \Delta W}{2 \omega_0^2}$$

$$C_1 = \frac{1}{\omega_0^2 L_1} - C_2 \left(1 - \frac{R_1}{R_2}\right)$$

$$C_2 = \frac{1}{\omega_0 \sqrt{R_1(R_2 - R_1)}}$$
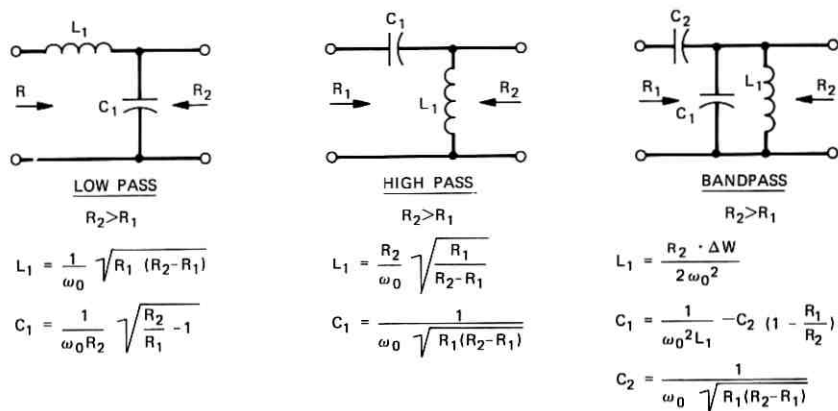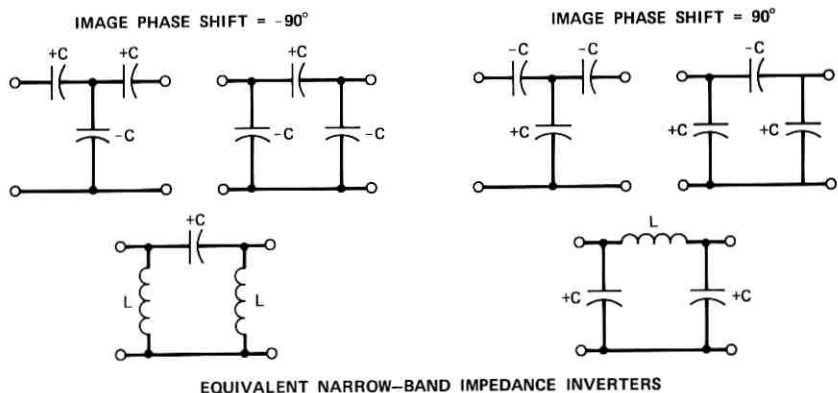
NARROW–BAND IMPEDANCE–MATCHING NETWORKS

Fig. 6—Narrow-band impedance inverters and impedance-matching networks.

resulting network can be identified as individual, two-resonator MCF's. Impedance inverters coupling the MCF's are also easily identified. If a composite filter design is sought, the impedance inverters indicated in the figure must be changed to their equivalent pi-configurations and the negative capacitance appearing in the series arms replaced by equivalent inductances on a narrow-band approximation basis. The final circuit arrangement is shown in the lower schematic of Fig. 7. It should be noted that, for the composite filter structure comprised of individual two-resonator MCF's, the design must be constrained to an even number of resonators.

The insertion-loss characteristic of a composite monolithic crystal bandpass filter, designed for use as a pick-off filter in the regulating repeater of the L5 system, is shown in Fig. 8.
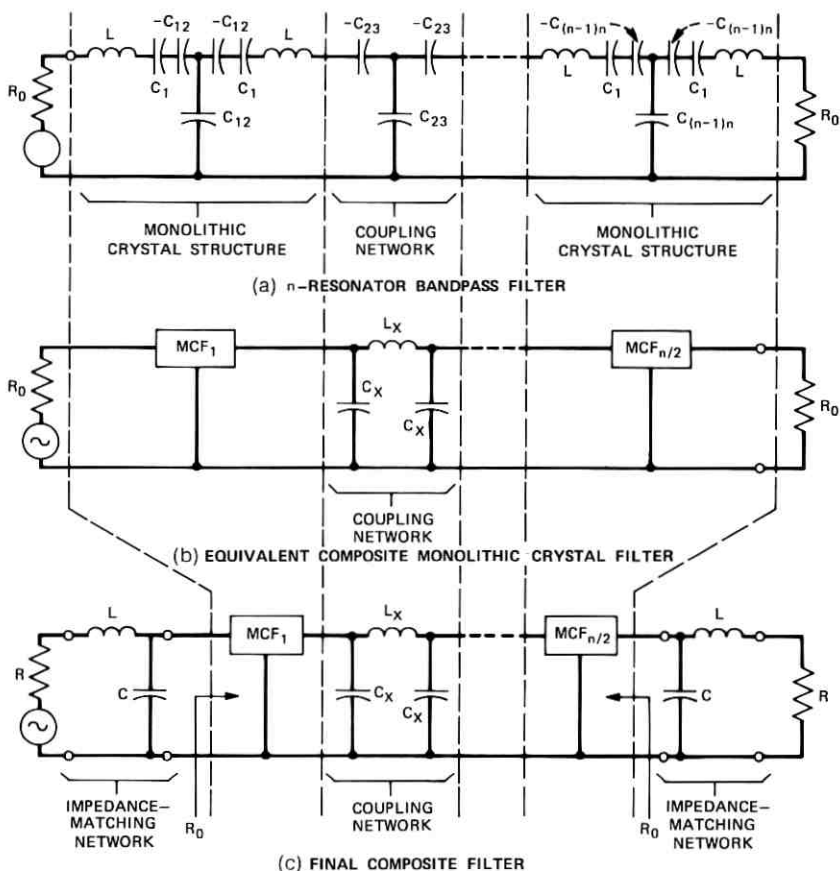
Fig. 7—Decomposition of an n-resonator bandpass filter to an equivalent, composite, monolithic crystal filter structure. (a) n-resonator bandpass filter. (b) Equivalent, composite, monolithic crystal filter. (c) Final composite filter.

1.5 Monolithic crystal band-elimination filters

Although the development of the monolithic crystal bandpass filter was first reported in early 1965, it was not until approximately three years later that the first disclosure of the monolithic band-elimination filter (BEF) was made. Two types of monolithic BEF[12] may be derived from monolithic crystal bandpass filters simply by the addition of inductive or capacitive elements connected across the ungrounded electrodes of conventional two-resonator monolithics. Different connections of the MCF plates, however, must be used to effect proper phase relations in the two types of filter. These phase relations are obtained by use of the basic or the reversed-phase connections previously
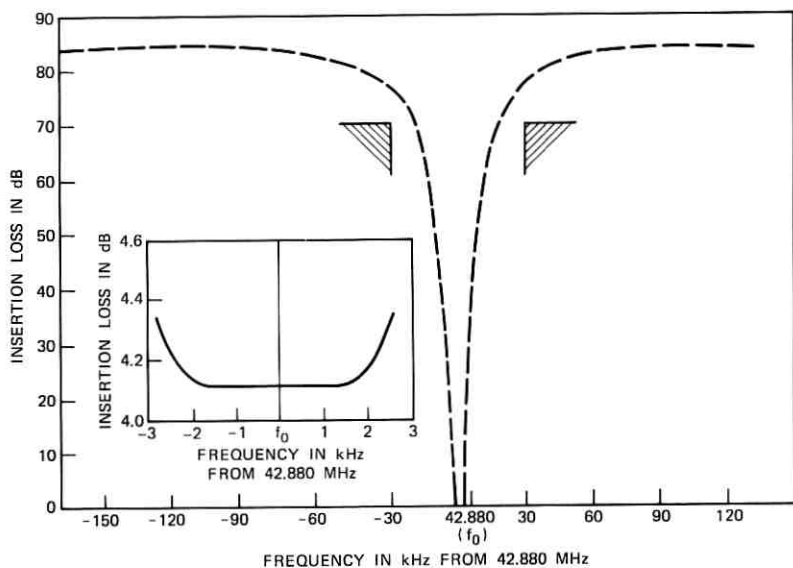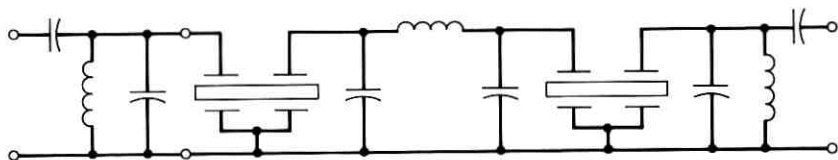
Fig. 8—Example of 42.880-MHz, monolithic crystal, bandpass filter.

shown for the bandpass filter of Fig. 5. Network schematics and equivalent circuits for the inductor- and capacitor-derived BEF's are shown in Fig. 9.

For the inductor-derived filter, the reject frequency occurs when the magnitude of the reactance of the bridging inductor equals that of the acoustic-coupling capacitors of the MCF. This reject frequency occurs at the center frequency of the bandpass filter. For the capacitor-derived filter, the reject frequency is obtained when the magnitude of the bridging capacitance is made equal to that of the acoustic-coupling capacitors.

The band-reject properties of either type of BEF are easily demonstrated by application of Bartlett's bisection theorem. The equivalent lattice of either network is first derived from the open and short-circuited impedances of the bisected network. Combining the redundant elements in the series and lattice arms of the lattice and

plotting reactance as a function of frequency, the passbands and stopbands can be readily identified. For the inductor-derived filter, the insertion loss characteristic includes a passband on either side of a reject band, with a second reject band extending to infinity beyond the second passband. The inductor-derived monolithic BEF is, therefore, a bandstop filter imbedded in a low-pass filter. The capacitor-derived filter, however, is a bandstop filter imbedded in a high-pass structure.

For practical applications, the inductor-derived filter exhibits superior passband performance compared to its capacitor-derived counterpart. Consequently, the remaining discussion of the monolithic BEF will be restricted to the inductor-derived circuit. The initial design procedures, used successfully on both single-section and multi-section monolithic BEF's, were based on dividing the frequency spectrum into two regions: (i) the stopband, over which the motional reactances of the MCF change rapidly as the crystal passes through its resonances, and (ii) the passband, over which the motional reactances of the MCF are sufficiently high to be considered as open circuits. The extreme stiffness of the resonators makes it possible, therefore, to attribute the frequency response, above and below the narrow stopband, solely to the low-pass filter section of the network. The narrow
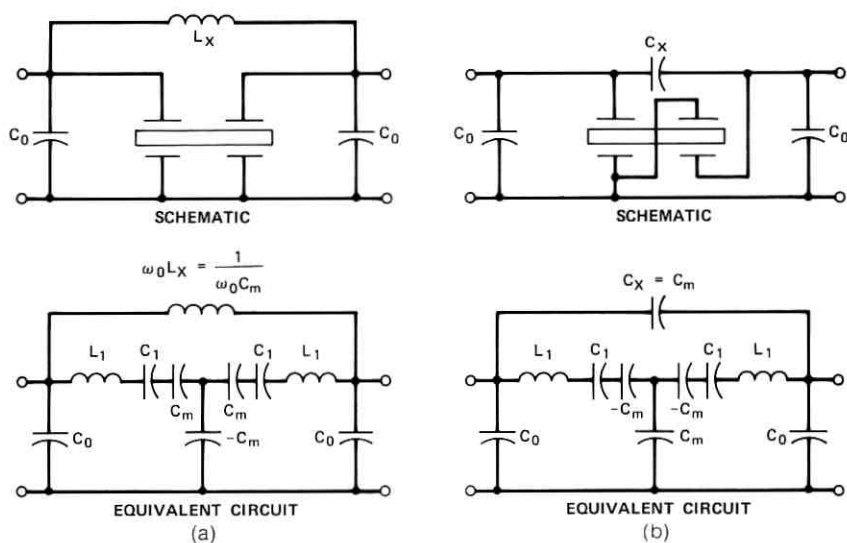

Fig. 9—Monolithic crystal, band-elimination filters. (a) Inductor-derived filter. (b) Capacitor-derived filter.

stopband, however, is controlled by the interactions of the MCF and the bridging inductor.

Single-section and multisection, inductor-derived, monolithic crystal BEF's, shown schematically in Fig. 10, may therefore be designed using standard low-pass synthesis techniques to meet passband requirements. The narrow reject band is formed by adding an estimated number of MCF's, with each monolithic bearing the proper relationship to the corresponding bridging inductor. The stopband performance can then be evaluated from an equivalent-circuit model of the entire network with crystal units being added or removed until the desired characteristic is obtained.

An alternative design technique to that outlined above results in a closed-form synthesis procedure in which a set of filter requirements such as passband ripple, cutoff frequency, and stopband bandwidth are related directly to the set of equivalent circuits shown in Fig. 11. This figure clearly delineates the low-pass and monolithic-crystal sections of the composite filter.
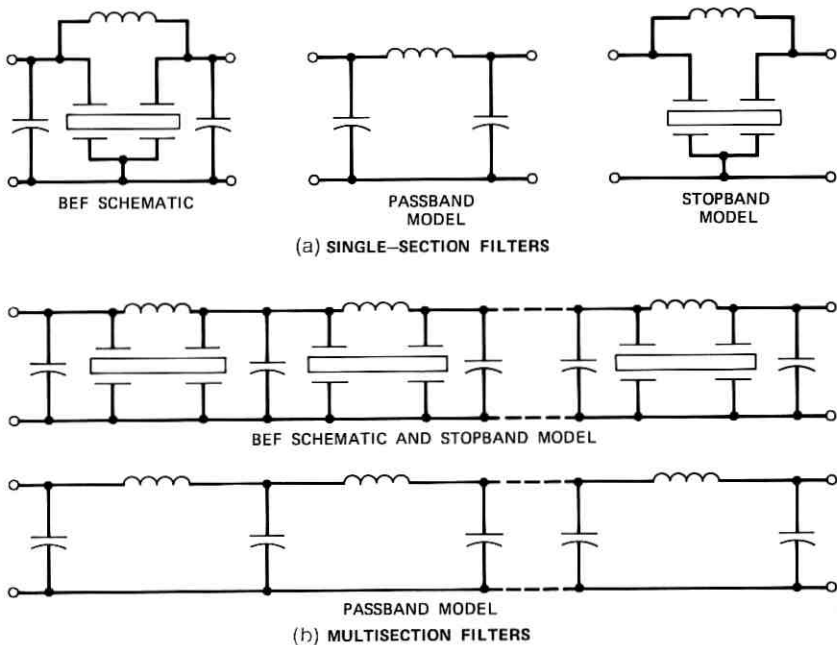

Fig. 10—Inductor-derived, monolithic crystal BEF's. (a) Single-section filters. (b) Multisection filters.

Writing the cascade matrix for each of the component networks of the composite structure, the complete matrix may be formed by converting each of the component matrices to its equivalent admittance form, adding the two resulting matrices and converting the sum back to the cascade form. Such a procedure is essentially routine and provides little insight into the synthesis procedure. If, however, these matrices are expressed in terms of dimensionless parameters[13] and a normalized frequency variable, the elements of a synthesis procedure become apparent. In each instance, the several dimensionless param-

EQUIVALENT CIRCUIT OF INDUCTOR–DERIVED, MONOLITHIC
CRYSTAL, BAND–ELIMINATION FILTER
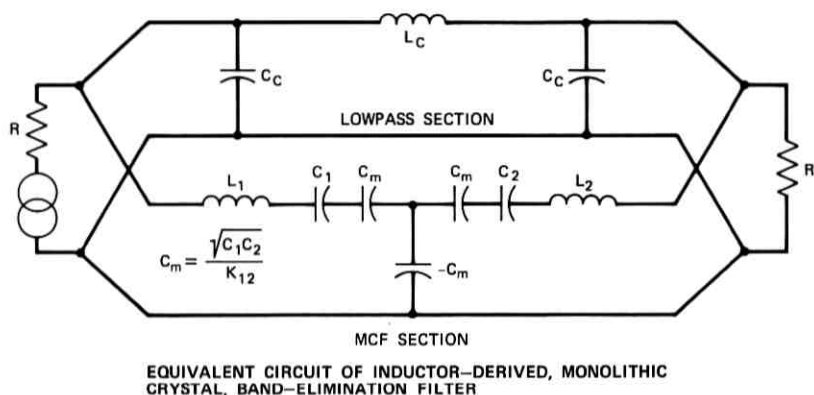
(a)

DIMENSIONLESS PARAMETER	DEFINING NETWORK SECTION
$\epsilon = 1 - \omega_0^2 L_C C_C$	LOWPASS FILTER
$D = R/\omega_0 L_C$	LOWPASS FILTER
$r = \omega_0^2 L_C C_m$	BEF $\left(\begin{array}{c}\text{COMBINED ELEMENTS}\\ \text{FROM LP AND MCF}\\ \text{SECTIONS}\end{array}\right)$
$\bar{L} = (L_1 + L_2)/\sqrt{L_1 L_2}$	MCF

NORMALIZED FREQUENCY

$$\Omega/K_{12}, \text{WHERE } \Omega = f/f_0 - f_0/f = 2(f - f_0)/f_0 *$$
$$\text{AND } K_{12} = 1/\left(\omega_0^2 \, C_m \, \sqrt{L_1 L_2}\right)$$

* NARROWBAND APPROXIMATION

DIMENSIONLESS PARAMETERS AND DEFINING NETWORK SECTIONS

(b)

Fig. 11—(a) Equivalent circuit of inductor-derived, monolithic crystal, band-elimination filter. (b) Dimensionless parameters and defining network sections.

eters are identified with the component low-pass and MCF's as shown in Fig. 11. The approach described has the further advantage of giving insight into previously unexplained asymmetries that occur in these designs. As a matter of interest, the cascade-matrix approach to network design has also been used to advantage in the design of mechanical filters.[14]

A complete discussion of the cascade matrix approach to the design of composite monolithic crystal BEF's is beyond the scope of the present paper; however, the usefulness of the method in describing seemingly anomalous behavior will be illustrated. A particularly interesting example is that of a two-resonator monolithic BEF having an asymmetric stopband and a Butterworth passband with selected parameters $\epsilon = 0.5$, $D = 1$ (see Fig. 11), and with $f_o < f_{co}$, where f_o is the reject frequency and f_{co} is the cutoff frequency of the low-pass filter. The resulting characteristic of this design is shown in Fig. 12. The solid-line plot is the insertion-loss or stopband model predicted on the basis of the foregoing development. The dashed-line characteristic is that of the Butterworth low-pass filter, and the circled points indicate the characteristic as derived from computer-based mesh analysis of the complete equivalent circuit. Two transition regions are to be noted: the first, below the reject frequency, f_o, and the second, above the reject frequency. In the region below the lower transition region and above the upper transition region, the circled points fall almost exactly on the computed characteristic of the low-pass filter by itself. In the region indicated as the stopband model region, the characteristic predicted by the cascade-matrix method of analysis is duplicated. In the transition regions, however, the approximation is not nearly as good. To simplify the analysis, it was assumed that the reactances of the low-pass filter remained constant in a narrow-band approximation. This approximation breaks down, however, over the transition and passband regions of the BEF. The results of the mesh-analysis computer runs, nevertheless, agree with the predicted characteristic in the stopband to at least three significant figures. The smoothest transition regions are attained in symmetric stopband designs when the reject frequency is placed at an insertion loss zero of a Chebychev low-pass filter. This is illustrated in Fig. 13 for the case of a 0.1579-dB ripple Chebychev passband.

Requirements for some of the BEF's specified for the L5 system included both high-loss stopbands (50 to 80 dB) and very wide passbands (approximately 70 MHz), coupled with low-distortion and high-return-loss requirements. Using Chebychev low-pass filters to
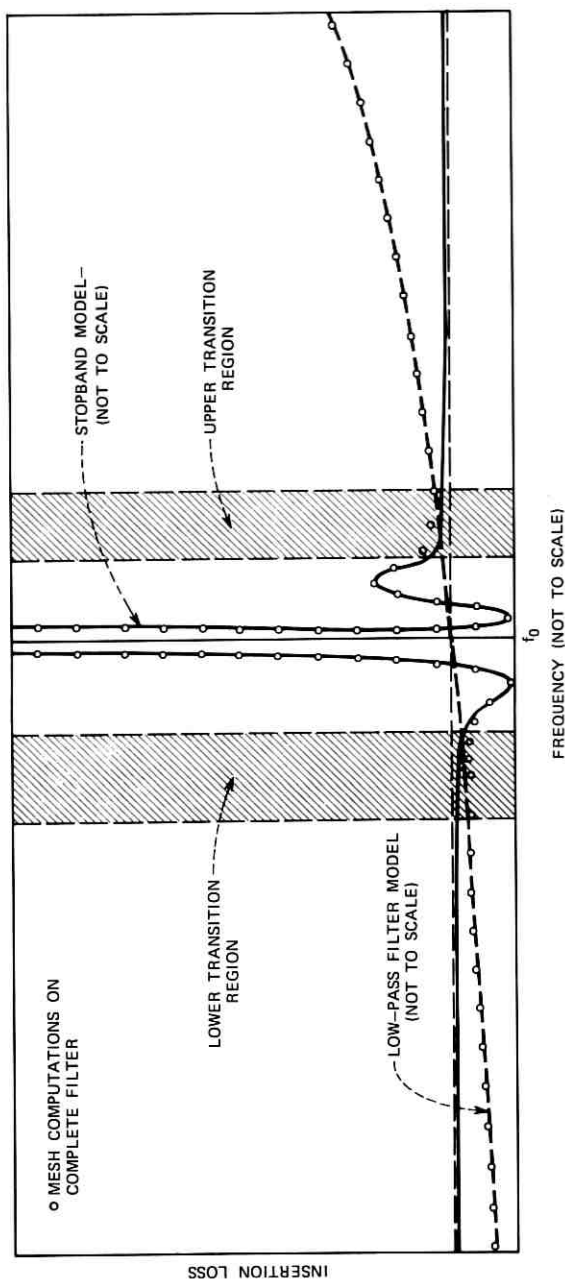
Fig. 12—Characteristic of band-elimination filter with asymmetric stopband and Butterworth passband. Transition regions are indicated ($\epsilon = 0.5$, $D = 1$, $f_o < f_{co}$).
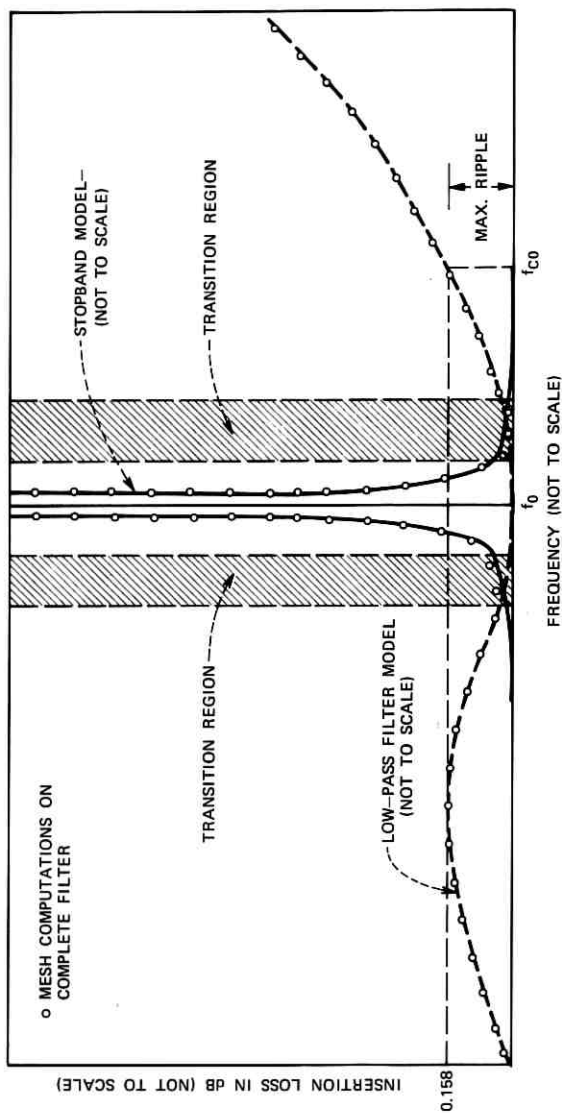
Fig. 13—Characteristic of band-elimination filter with symmetric stopband and 0.1579-dB-ripple Chebychev passband. Transition regions are indicated. Stopband center frequency f_o is at insertion loss zero of the low-pass filter.

model the low-pass section of the composite filter, it was found possible to meet all systems requirements. A specific example of design capability is given in Fig. 14 for a 42.880-MHz BEF, using three two-resonator monolithics imbedded in a ninth-order Chebychev low-pass filter. This low-pass filter was designed at an impedance level of 450 ohms to achieve optimal inductance values for the MCF's. The filter was then matched to the 75-ohm source and load impedances by use of auto-transformers, as shown in the figure. Impedance-matching sections, previously described for bandpass designs, could not be used because of the narrow-band approximations involved. To meet the passband insertion-loss deviation requirements of ±0.05 dB from flat response over the entire frequency range of the L5 system, a simple bridged-T amplitude equalizer was also required.

NETWORK SCHEMATIC

Fig. 14—Example of 42.880-MHz pilot-blocking filter.

1.6 Adjustable Bode equalizers[15,16]

In the L5 coaxial system, a total of 28 adjustable bump shapes are provided to correct for time-invariant amplitude distortions.[17] Ten of these shapes are allocated to the E1 equalizer for coarse correction of distortions introduced by variations in the manufactured product. The remaining 18 shapes are provided by the E2 equalizer to correct whatever residual distortions remain after employment of all lower levels of equalization, including the E1 equalizer.

The majority of bump shapes are realized by series-type, multi-bump, adjustable Bode equalizers. The remainder are supplied by shaping networks introduced into the feedback paths of isolating amplifiers. The amplifiers buffer the input and output ports of each Bode network and supply gain to offset the losses introduced by the equalizers. Each of the multibump equalizers, however, introduces no more flat loss than would have been introduced by any one of the several single-bump equalizers required to duplicate its performance. With a reduction in the number of Bode networks, the number of isolating amplifiers is also reduced. This, in turn, results in a reduction in inter-modulation and thermal noise. Finally, the introduction of shaping networks into the feedback of the isolating amplifiers also reduces the number of Bode networks required. The net effect of all these consider-ations is to improve reliability, to reduce cost, and as previously in-dicated, to reduce intermodulation and thermal noise.

All of the multibump Bode networks supplied for the E1 and E2 equalizers were derived from extensions of the simple series-type adjustable network of Fig. 15a. The basic shaping network, as indicated in the figure, is the bridged-T equalizer. Figure 15b shows the range of insertion-loss shapes that may be obtained by variation of load resistor, R_a.

As demonstrated by Lundry,[18] the insertion loss of the generalized Bode network of Fig. 16a can, to a good approximation, be expressed as

$$\theta - \theta_o = \kappa \cdot \rho_a \epsilon^{-2\phi}, \tag{1}$$

where

ϕ = the image transfer constant of the bridged-T network

θ = the insertion loss of the adjustable equalizer

θ_o = the insertion loss of the equalizer when $R_a = R_o$

$\kappa = 2 \tanh \dfrac{\theta_o}{2} = \dfrac{\epsilon^{\theta_o} - 1}{\epsilon^{\theta_o} + 1}$

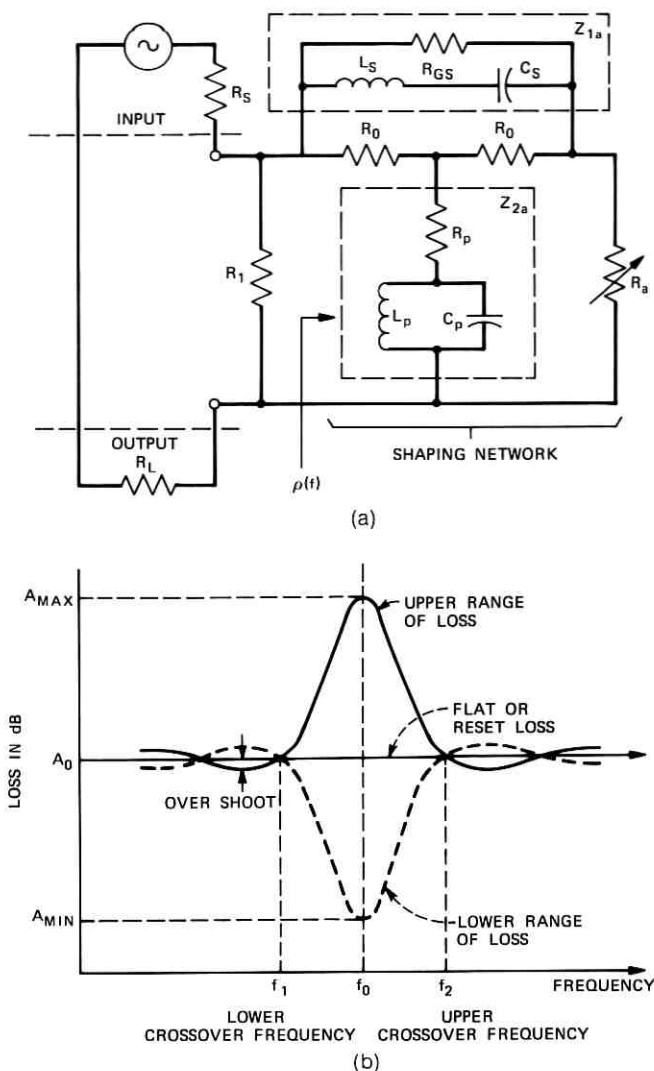ρ_a = reflection coefficient of R_a against R_o.

Fig. 15—Series-type, single-bump, adjustable, Bode equalizer. (a) Schematic. (b) Characteristic.

Equation (1) can also be expressed as

$$\theta - \theta_o = \kappa \cdot \rho(f), \tag{2}$$

where $\rho(f)$ is the input reflection coefficient of the shaping network relative to R_o and is a function of frequency. Considering only the real
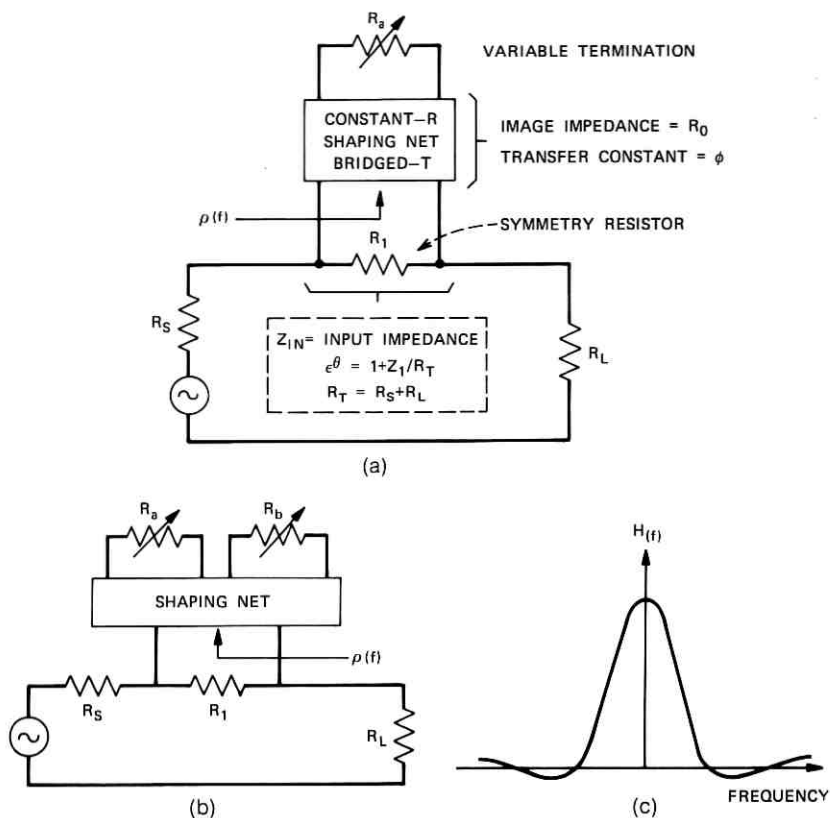
Fig. 16—(a) Series-type, single-bump, Bode equalizer. (b) Series-type, double-bump, Bode equalizer. (c) Bump characteristic, $H(f)$.

part of eq. (2), the insertion loss can be written as

$$\alpha(\text{dB}) = \alpha_o + \text{Re}\,[\kappa \cdot \rho(f)], \tag{3}$$

where α_o is the insertion loss of the structure when $R_a = R_o$. Since α_o and κ are constants, it is only necessary to consider $\rho(f)$.

To realize a single-bump shape with the network of Fig. 16a, $\text{Re}\,|\rho(f)|$ is required to have the form

$$\text{Re}\,|\rho(f)| = H(f)\rho'(R_a), \tag{4}$$

where $H(f)$ is a bump-shaped function of frequency (Fig. 16c), and $\rho'(R_a)$ is a function of a single resistor, independent of frequency.

Similarly, to realize a double-bump shape with the network of Fig. 16b, $\rho(f)$ will have the form

$$\text{Re } |\rho(f)| = H_a(f)\rho_a'(R_a) + H_b(f)\rho_b'(R_b), \tag{5}$$

where the subscripts a and b refer to two separate bump shapes. Both $H_a(f)$ and $H_b(f)$ have the shape indicated in Fig. 16c. For the type of designs discussed here, it is required that these shapes be in distinct frequency bands. R_a and R_b provide independent control of the two shapes.

One method of realizing eq. (5) is to use bridged-T sections as shaping networks. To demonstrate this, a general expression for the input reflection coefficient of a bridged-T network is required. Referring to the schematic of Fig. 17b, which is an unconventional arrangement of the circuit of Fig. 17a, the scattering matrix[19] of the network, with $Z_1 \cdot Z_2 = R_0^2$, is given by

$$\begin{bmatrix} b_1 \\ b_2 \\ b_3 \\ b_4 \end{bmatrix} = \begin{bmatrix} 0 & S_{12}(f) & 0 & S_{14}(f) \\ S_{12}(f) & 0 & S_{14}(f) & 0 \\ 0 & S_{14}(f) & 0 & S_{12}(f) \\ S_{14}(f) & 0 & S_{12}(f) & 0 \end{bmatrix} \cdot \begin{bmatrix} a_1 \\ a_2 \\ a_3 \\ a_4 \end{bmatrix}, \tag{6}$$

where a_i and b_i refer to the incident and reflected voltage waves at the ith port, respectively, and

$$S_{12} = (1 + Z_2/R_o)^{-1}$$
$$S_{14} = (1 + Z_1/R_o)^{-1}.$$

The reflection coefficient $\rho(f) = b_1/a_1$, looking into port 1 of the network, must now be determined. Ports 2 and 4 are terminated in impedances whose reflection coefficients are $\rho_2(f)$ and $\rho_4(f)$, respectively. Assume port 3 is terminated in R_o, and as a result, $\rho_3(f) = 0$.
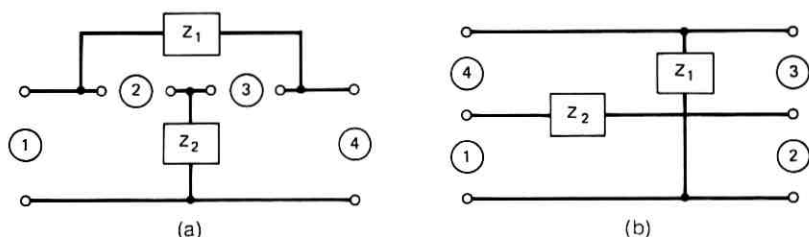

Fig. 17—Bridged-T networks with ports 1 through 4 open-circuited. (a) Conventional. (b) Unconventional.

Thus, the port relationships are

$$a_2 = \rho_2(f)b_2$$
$$a_4 = \rho_4(f)b_4 \qquad (7)$$
$$a_3 = 0.$$

These port relationships, together with eq. (6), lead to

$$\rho(f) = S_{12}^2(f) \cdot \rho_2(f) + S_{14}^2(f) \cdot \rho_4(f). \qquad (8)$$

Terminating port 2 in R_o and port 4 in R_a leads to

$$\rho(f) = S_{14}^2(f)\rho_4(R_a),$$

where $\rho_4(R_a)$ has been substituted for $\rho_4(f)$, since a resistive element has replaced a frequency-dependent impedance. This has the form of the reflection coefficient for a single-bump shape defined by eq. (4). The single-bump network, based on the unconventional bridged-T network of Fig. 17b, is shown in Fig. 18a.

Because $Z_1 \cdot Z_2 = R_o^2$, $S_{12}(f)$ and $S_{14}(f)$ are related by

$$S_{14}(f) + S_{12}(f) = 1. \qquad (9)$$

Assume that the shaping network is such that $\mathrm{Re}\,[S_{14}^2(f)]$ is a bump shape centered at frequency f_a. As a result, $S_{14} \approx 1$ in the vicinity of f_a and $s_{14} \approx 0$ elsewhere. From eqs. (8) and (9)

$$\rho(f) = \begin{cases} S_{14}^2(f) \cdot \rho_4(R_a) & \text{(in the vicinity of } f_a) \\ S_{12}^2(f) \cdot \rho_2(f) & \text{(elsewhere).} \end{cases}$$

In this way, ports 2 and 4 are decoupled.

The term $S_{14}^2(f) \cdot \rho_4(R_a)$ represents a bump shape. As a result, half of the desired equation, (5), has been realized. To realize the second bump shape, $\rho_2(f)$ is made the input reflection coefficient of a similar shaping network whose center frequency is f_b, with $f_b \ll f_a$ or $f_b \gg f_a$. Networks having the assumed properties for S_{14} are shown for a two-bump equalizer in Fig. 18b. For this network,

$$\rho(f) = S_{12a}^2(f)S_{14b}^2(f)\rho_{4b}(R_b) + S_{14a}^2(f)\rho_{4a}(R_a). \qquad (10)$$

Equation (10) is now in the proper form to realize the defining equation, (5), of the double-bump equalizer. The equivalence of the two equations becomes apparent when the following substitutions are made in eq. (10):

$$\mathrm{Re}\,[S_{12a}^2(f) \cdot S_{14b}^2(f)]\rho_{4b}(R_b) = H_b(f)\rho_b'(R_b)$$

and

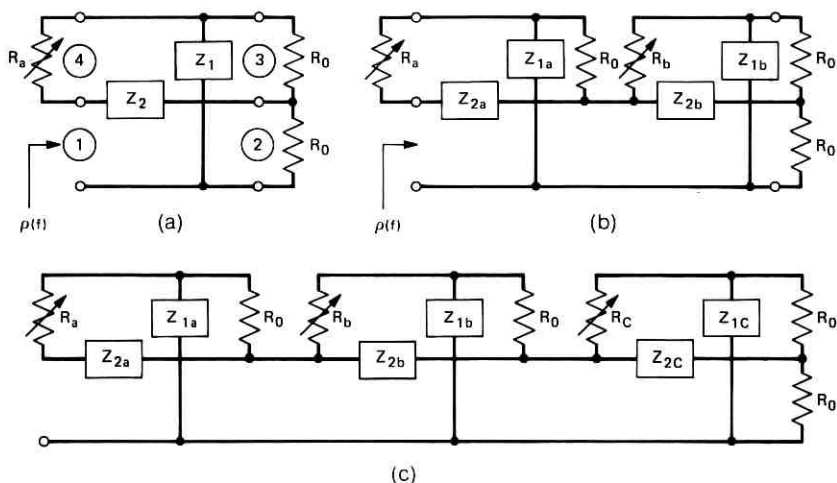$$\mathrm{Re}\,[S_{14a}^2(f)]\rho_{4a}(R_a) = H_a(f)\rho_a'(R_a).$$

Fig. 18—Shaping network for adjustable bridged-T Bode equalizers. (a) Single bump. (b) Double bump. (c) Triple bump.

The technique described above can be easily extended to the realization of "*n*" independent bump shapes. The method becomes more or less obvious when the circuit diagrams of Fig. 18a, b, and c, for single-, double-, and triple-bump equalizers are compared.

1.7 Computer-aided network tuning applied to $\sqrt{f}$ regulating network

One of the more recent applications of digital computers is computer-aided tuning of complex networks in manufacture. Without on-line computer adjustment, the $\sqrt{f}$ regulating network of the L5 system could not have been manufactured without relaxation of the performance criteria by a factor of several times. COTMS provided an ideal facility for implementing the alignment algorithm described here.

To compensate for changes in cable loss relative to ground-temperature variations, regulating repeaters are placed in the L5 line approximately every 5 to 7 miles.[17] Relative to nominal cable loss at mean ground temperature, both gain and loss compensation are required according to whether the temperature is above or below mean temperature. In either case, the shape, on a dB (log) basis, is proportional to the square root of frequency. The networks that provide the compensation, as in the L4 system,[20] are thermistor- and pilot-controlled, wideband, square-root-of-frequency Bode networks. They differ from the L4 designs, however, in that buffering amplifiers are included as integral components of the networks, as shown in Fig. 19. Design
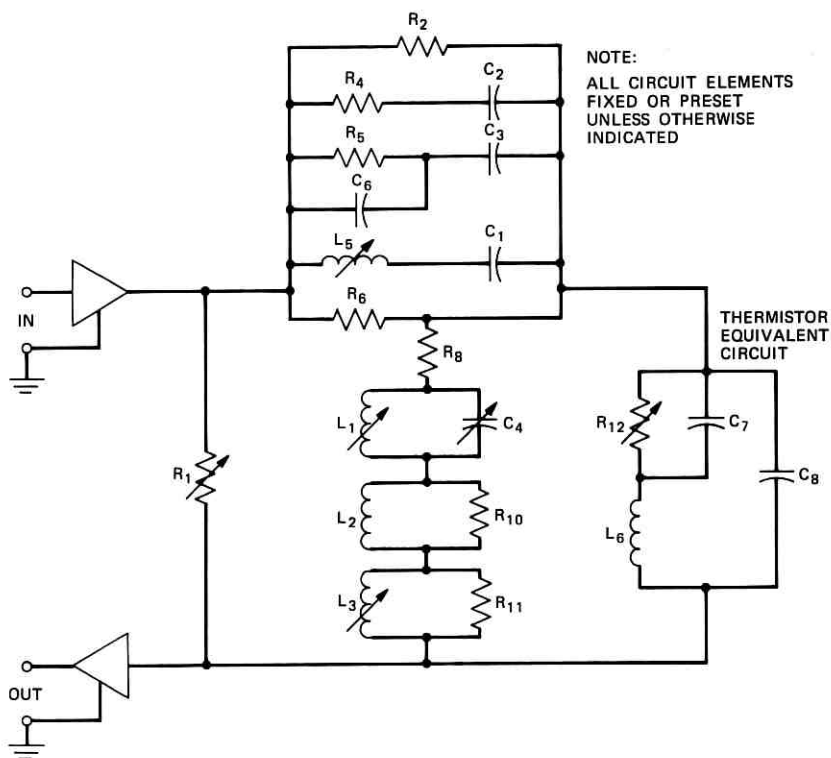
Fig. 19—Simplified schematic of square-root-of-frequency Bode network.

considerations, however, are basically the same and are not reconsidered.

Three measurements are required to fully characterize the network: (i) at high-loss setting, (ii) at nominal setting, and (iii) at low-loss setting (see Fig. 20). The dynamic errors may be defined as

$$e_H(f) = A_H(f) - A_N(f) - K_H\sqrt{f}\ \mathrm{dB} \qquad (11)$$

$$e_L(f) = A_L(f) - A_N(f) + K_L\sqrt{f}\ \mathrm{dB}, \qquad (12)$$

where $A_H(f)$, $A_N(f)$, and $A_L(f)$ are loss measurements in the high-, nominal-, and low-loss conditions, respectively, and K_H and K_L are determined by requiring the error to be zero, by definition, at the pilot frequency of 42.88 MHz. The static error may be defined as

$$e_N = A_N(f) - A_R(f)\ \mathrm{dB}.$$

$A_R(f)$ is the expected loss in the nominal condition, which, because of parasitics, gain shaping in the amplifiers, and interaction effects, is not flat with frequency. Deviations from the ideal flat shape, however, have the same effect as deviations from the nominal characteristic of other components in the transmission path and may be corrected by fixed and adjustable equalizers available in the system. Since comparable dynamic equalizers are not available, more stringent requirements are placed on the dynamic performance of the Bode equalizer than on the flat or reset shape.

The necessity of requiring three loss measurements, corresponding to three thermistor settings, to characterize network performance, precludes the use of conventional methods for compensating parasitic variations in the amplifiers and the network during manufacture. As it is not possible to guarantee network performance from easily measured amplifier parameters, the alignment procedure must correct for both network and amplifier variations. Since the amplifier is not adjustable, the network must provide tuning capabilities for both shaping and amplifier compensation. Simple prescription of element
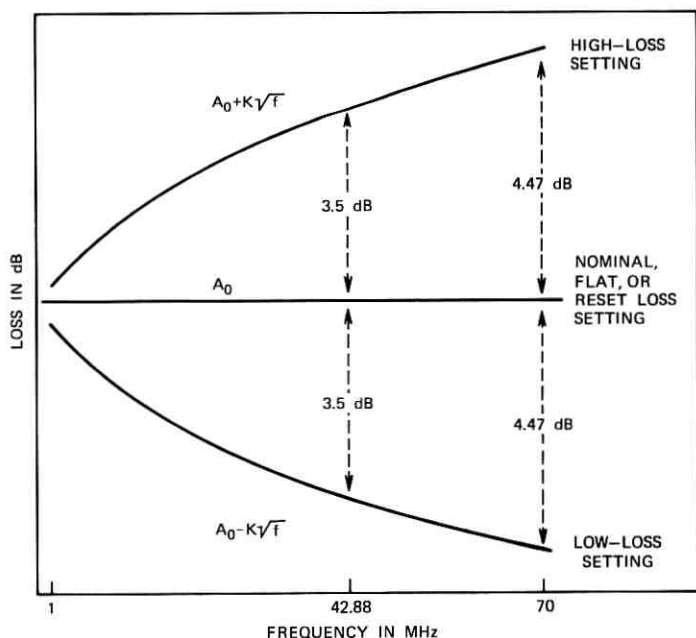

Fig. 20—Square-root-of-frequency Bode-network characteristic.

values or adjustment of resonances would not provide this capability. Furthermore, an adjustment in a network component usually affects all three shapes, making adjustment on a scanner too difficult. Thus, a more sophisticated procedure is needed.

Since the amount of adjustment needed is small, but the allowed deviations even smaller, linearization has proved to be valid. The error after adjustment, $e_{HA}(f)$, can therefore be expressed as:

$$e_{HA}(f) = e_{HI}(f) + \sum_j x_j \Delta_{HJ}(f) \text{ dB}, \qquad (13)$$

where $e_{HI}(f)$ is the initial error, $\Delta_{HJ}(f)$ is the normalized deviation caused by the jth component, and x_j is the amount of the deviation used for adjustment. Similar relations hold for the nominal- and low-loss errors. The composite error may be defined as the summed weighted squares:

$$e^2 = \sum_{\text{all} f} [e_H^2(f) + e_L^2(f) + W e_N^2(f)]. \qquad (14)$$

A weighting W less than unity allows a larger error in the nominal or reset shape than in the dynamic errors e_H or e_L. Given an initially determined set of errors for the three conditions, over an appropriate frequency band, a set of parameter deviations x_j may be found that minimizes e^2, using the so-called least-sum-of-squares procedure. This procedure is well known and only requires the solution of a set of simultaneous linear equations.

Unfortunately, in practice, the unconstrained solution often requires an adjustment greater than the adjustable components can provide. Thus, it was necessary to include a search for the best constrained solution. The method chosen uses gradient information to determine if a parameter must violate the constraints to lessen the summed-squares error, e^2. Other than being iterative, the method does not require any information beyond that obtained from the least-sum-of-squares procedure. While inclusion of the constraint algorithm contributed to the success of the program in manufacture, a complete description is omitted because of space limitations. It is necessary, of course, to preset the adjustments to a prescribed condition, prior to alignment, to correspond to bounds stored in the program.

Programming for the constrained least-sum-of-squares algorithm was added to a streamlined measurement program for COTMS. The computer controls the measurement sequence and the alignment procedure automatically. The only manual tasks required during tuning are to switch the network into the different conditions and to turn the slugs on the adjustable components. Measured, rather than
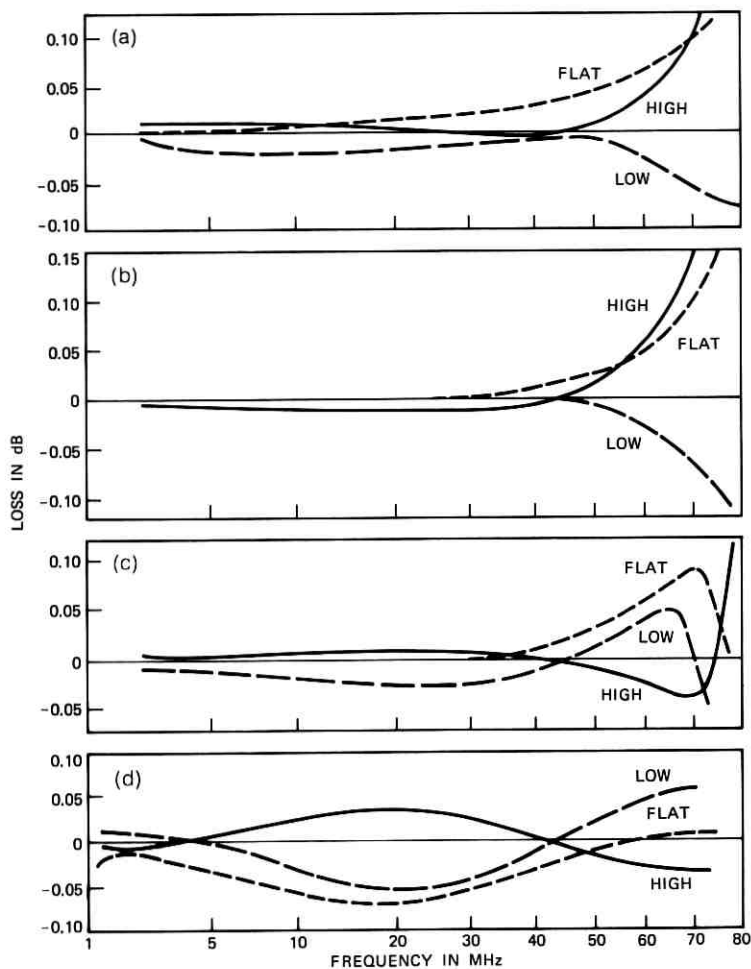
Fig. 21—Deviations in error with element adjustment in square-root-of-frequency Bode network. (a) L1. (b) C4. (c) L5. (d) L3.

computed, deviations from a typical network are used due to the parasitic modeling problem. Because adjustments are effected on a loss basis, it is neither necessary to count turns of the adjustable inductor slugs nor necessary to measure inductance.

The deviations in the high-, nominal-, and low-loss shape errors, caused by variations of four components, are shown in Fig. 21. The curves for inductor L1, for example, show the deviation in the errors corresponding to a 0.1-dB change at 70 MHz in the nominal- or flat-

loss-setting response. The result of the alignment algorithm is a percentage of this deviation. For example, if the multiplier x_j of eq. (13) is determined by the least-sum-of-squares procedure to be 1.32, multiplication of inductor L1 sensitivity (0.1 dB) by this number results in 0.132 dB. Inductor L1 must, therefore, be adjusted for a 0.132-dB change in loss, at 70 MHz, to effect the optimum tuning of that element. For inductor L3, the maximum deviations from flat response occur at approximately 20 MHz over the 1- to 70-MHz frequency range. This inductor is, therefore, tuned at that frequency.

Six adjustable components are included in the program. Only one measurement, adjustment, and remeasurement iteration proved necessary for the adjustment of the networks for optimum performance. "Before" and "after" adjustment error curves are shown in Fig. 22.
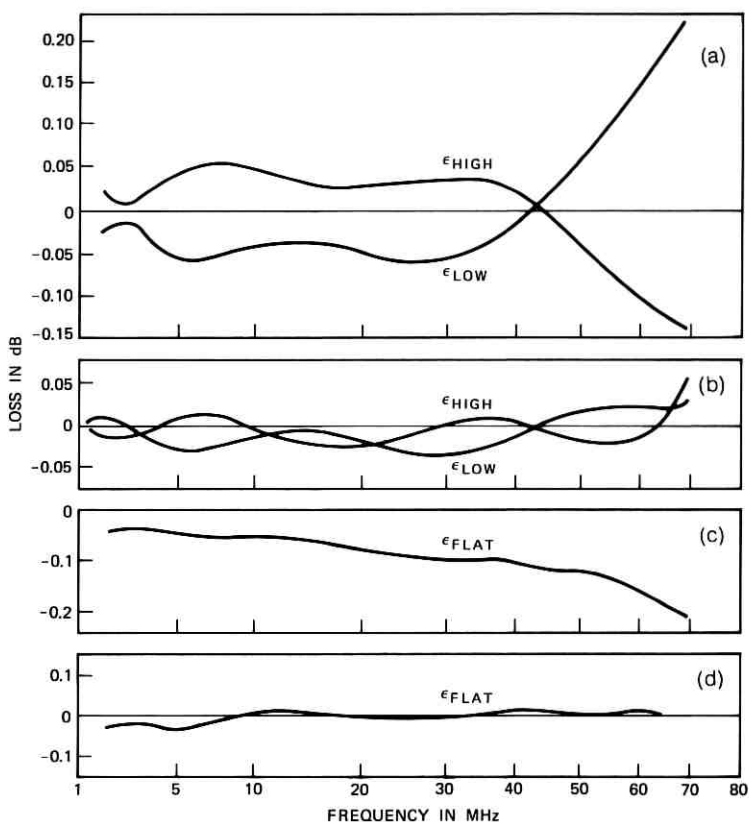

Fig. 22—Network errors before and after adjustment. (a) Before adjustment. (b) After adjustment. (c) Before adjustment. (d) After adjustment.

The computer-aided-adjustment program has proved its value during the several years of manufacture of the regulating network. Without COTMS, this approach would not be feasible. Without computer-aided adjustment, the dynamic regulating objectives for the L5 system could not have been met.

1.8 Deviation equalizers[17]

In a system having as complex an equalization arrangement as L5, the possibility of advantageous interactions among equalizers in the complete hierarchy of equalization must be considered in any one design. An example of such design considerations is the design of the relatively simple, fixed-loss, deviation equalizer. This equalizer compensates for the average error in the match of average-line-repeater gain to nominal cable loss. Three deviation equalizers are installed in each power-feed section.

Initial evaluation of the design led to the conclusion that requirements could be met with an equalizer consisting of two valley-shaped and two bump-shaped bridged-T equalizer sections connected in tandem, as shown in Fig. 23. It was necessary, however, that the design be optimized so that the equalizer, in combination with the E1 and E2 adjustable networks, would provide optimum compensation. This was achieved using the general-purpose optimization program (GPOP) tied together with an E1, E2 simulation program.[21]

The E1, E2 program contains the measured response of all the bump shapes included in the E1 and E2 equalizers. With approximately 15
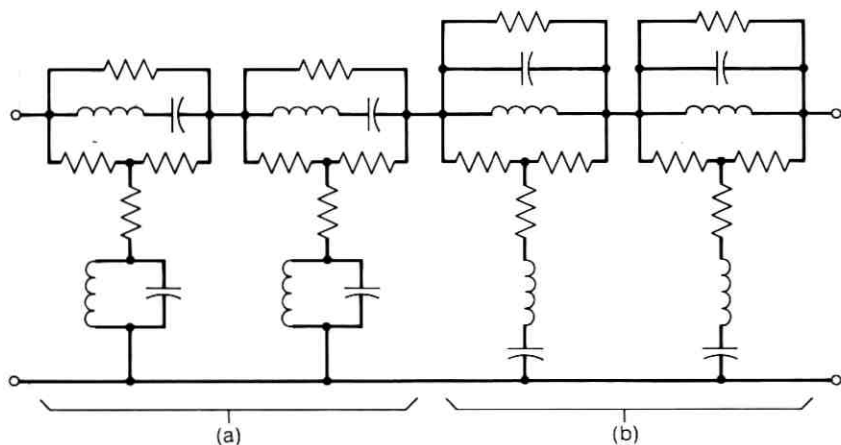

Fig. 23—Deviation equalizer. (a) Valley-equalizer sections. (b) Bump-equalizer sections.

percent of each E1 and E2 bump allotted to the deviation equalizer, optimization resulted in an equalizer characteristic having deviations from ideal placed at frequencies where the E1 and E2 adjustable equalizers could best supply shaping to reduce the overall misalignment.

The E1, E2 simulator package simulated the iterative process that a craftsperson would use in the field in making system equalization adjustments. This yielded an effective design which has led to excellent results on equalized lines.

1.9 Physical design

Although essentially every transmission network for the L5 system provided its own challenges in physical realization, the majority of challenges related to compensation for parasitic inductance and capacitance, to minimization of ground loops, and to reduction of coupling between components. Environmental protection was provided by drawn or fabricated metal enclosures with internally mounted PWB's for supporting and interconnecting the individual circuit components. External connections were, in most instances, made by use of moisture-resistant plugs or jacks.

In several instances, however, substantially greater physical-design effort was involved. Two examples of such design effort are: (i) the design of the shaping networks of the line repeaters, and (ii) the physical design of the earth-ground filter.

The shaping networks of the basic, regulating, and equalizing repeaters were specifically designed to reduce excess low-frequency gain, to furnish surge protection, and to provide the power-separation filtering requirements of the L5 repeaters. Figures 24a and b show the schematics of the low-frequency A and B networks. The circuit diagrams identify those portions of the networks concerned with the different circuit functions. The network schematic of the bridged-T equalizers used to obtain the desired insertion loss is shown in Fig. 24c.

The physical design of the shaping networks required considerable interaction with system engineers. For overall efficiency and economy of space utilization, the surge-protection and power-separation circuits were included in the shaping-network package. As indicated in the photograph of Fig. 25, the two networks were placed side by side in the lower section of the basic repeater housing. Requirements on RF isolation, between input and output ports of the repeaters, led to the use of cast-aluminum housings for packaging the shaping networks. These housings, in turn, furnished the necessary mechanical
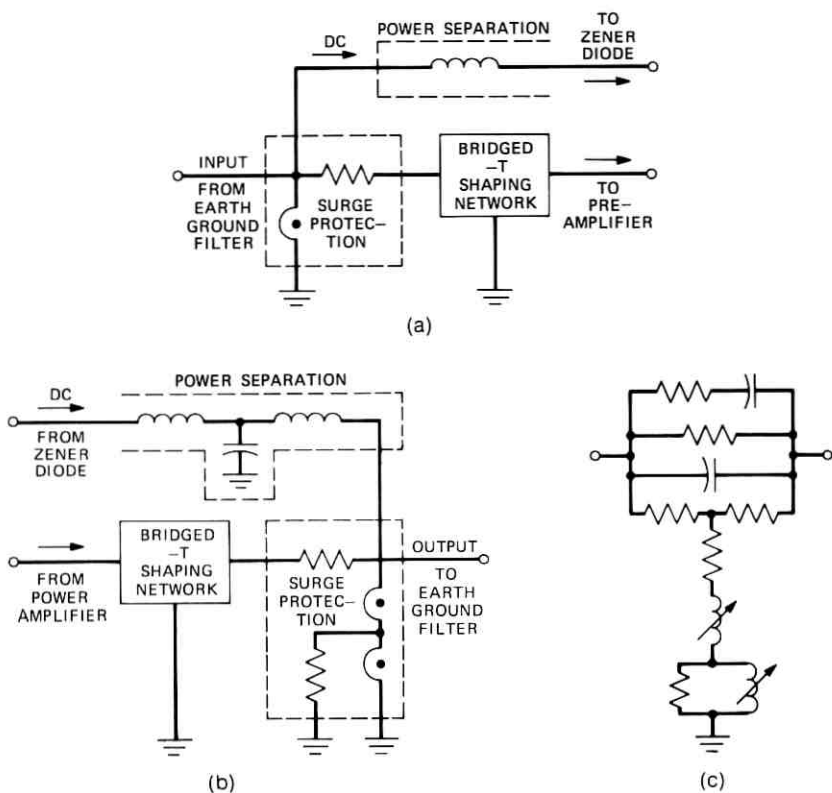
Fig. 24—Low-frequency shaping networks. (a) Network A. (b) Network B. (c) Bridged-T network schematic.

rigidity to meet the ± 0.025-dB tolerance on insertion-loss shape as specified for manufacture.

The earth-ground filters of the line repeaters, shown in the photograph of Fig. 25, with their coaxial jacks protruding from the repeater housing, perform several important functions. Despite the seemingly simple electrical circuit shown at the input and output ports of the simplified block diagram of the basic repeater of Fig. 26, the earth-ground filter must

(*i*) Provide high-voltage dc blocking between earth and repeater ground.

(*ii*) Maintain a low-impedance return path between the grounds for all frequencies in the L5 band.
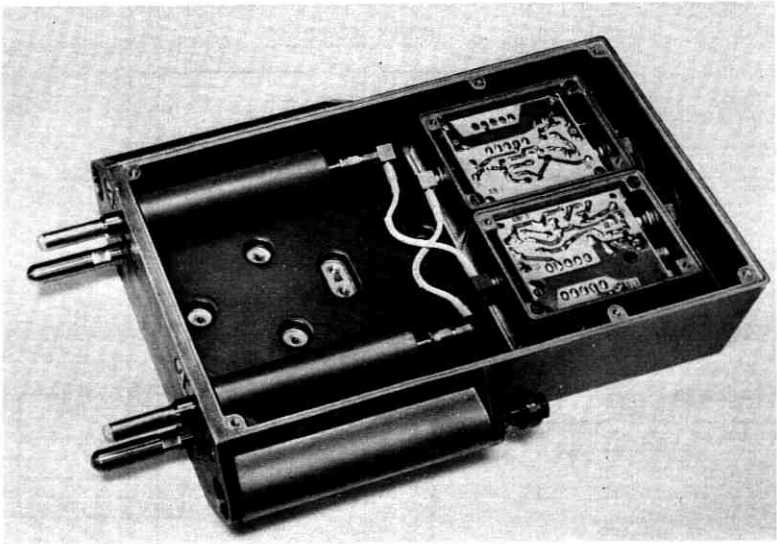
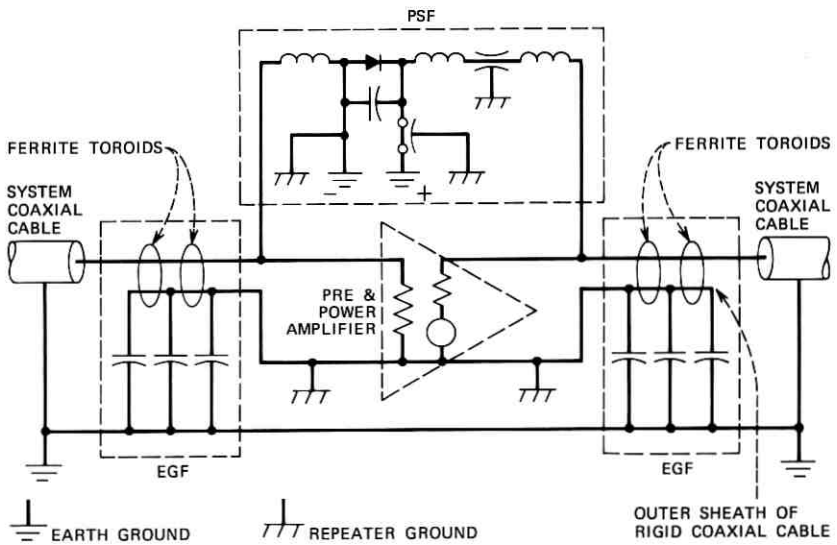Fig. 25—Earth-ground filters and shaping networks in repeater housing.

Fig. 26—Block diagram of L5 repeater.

(*iii*) Provide a low-pass filter characteristic with a minimum of 120-dB isolation between repeater input and output (through the ground return) for signals in the L5 band.

(*iv*) Carry the L5 line frequencies and the dc powering current with a minimum of amplitude distortion.

The physical design of the earth-ground filter[22] presented a design challenge. As shown in Fig. 27, the filter contains two coaxial chokes in the form of ferrite toroids and three high-voltage, high-reliability, dual-dielectric capacitors packaged in a sealed enclosure. The filter, minus the coaxial plug, is impregnated with polyisobutylene and hermetically sealed to prevent the entrance of moisture or the leakage of impregnant, which would contaminate the connectors.

To insure the high reliability of this design, considerable emphasis was placed on the compatibility of materials, structural integrity, sealing, and manufacturing control and testing.

II. MAGNETIC COMPONENTS

2.1 Introduction

The L5 system required the development of a wide variety of magnetic components numbering in excess of 100 different experimental designs. This number includes original designs, subsequent improved versions, and redesigns that reflected changing circuit concepts or requirements. Ultimately, about 50 apparatus codes were issued to specify these designs for manufacture. Many additions to existing code series of magnetic components were also made available for L5 applications, but these are not discussed. The adjustable, air-core, precision inductors specified for the manufacture of many of the transmission networks developed for L5 are an example of magnetic component designs based on existing code series.

Coded transformers and inductors appear in all segments of the L5 repeatered line. Frequently, these components are associated with thin-film circuits and, as a result, miniaturized designs had to be made available. Development of inductor structures for such applications, however, predated the development of the L5 system, and, as a result, emphasis is placed here on the development of various types of transformers. Many transformer designs for the line repeaters required pin-type terminals for interconnecting and mounting on PWB's or thin-film substrates. These transformers are discussed in Section 2.2, Repeater Components.
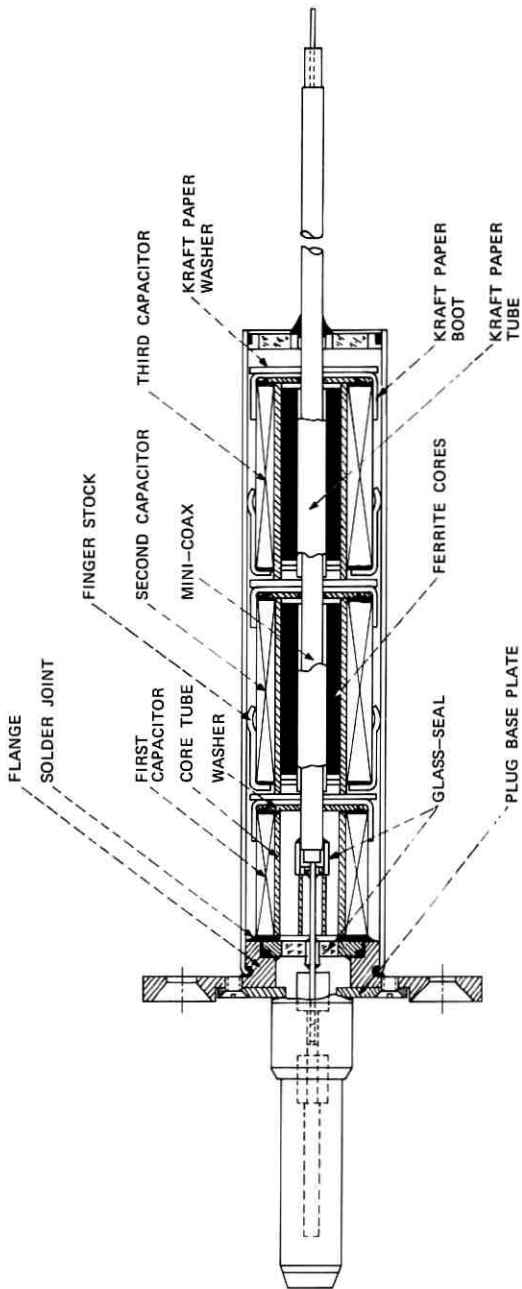
Fig. 27—Cross-sectional view of the earth-ground filter.

Labels (top to bottom / left to right):
KRAFT PAPER WASHER
THIRD CAPACITOR
KRAFT PAPER BOOT
KRAFT PAPER TUBE
SECOND CAPACITOR
MINI—COAX
FINGER STOCK
FERRITE CORES
FLANGE
SOLDER JOINT
FIRST CAPACITOR
CORE TUBE
WASHER
GLASS-SEAL
PLUG BASE PLATE

Magnetic components intended for terminal applications are frequently used in complex arrays to develop hybrid trees arranged for coaxial interconnection. The different methods used to establish an interface between the magnetic components and the circuitry of the system reflected not only on the problem of physical design, but on the problem of electrical realization as well. Requirements on the individual transformers constituting the hybrid trees had to be tightened to meet overall system performance criteria. As a result, these magnetic components had to be built out with resistors and capacitors to form what are, in effect, special types of transmission networks. Although this type of component is also used to a limited extent in the regulating and equalizing repeaters, a representative design is described in Section 2.3, Terminal Components. Finally, since measurement limitations and techniques were important to all classes of designs, these are discussed separately in Section 2.4.

2.2 Repeater components[17]

The amplifiers in the basic and regulating line repeaters make extensive use of bridge-type feedback,[23] which offers the advantage of making the $\mu\beta$ loop independent of the line impedance Ordinarily, the disadvantages of bridge-type feedback are the extreme impedance levels and the power consumed in the bridge arms used to secure a bridge balance. Both of these limitations, however, may be overcome by use of hybrid transformers. Because of the very tight limits, typically ± 0.02 dB, placed on reproducibility over the 1- to 70-MHz passband, transmission-line design techniques[24] were extensively adapted to the design of hybrid transformers. These designs specify pairs, triplets, or quadruplets of insulated magnet wire twisted uniformly together and wound around a ferrite core. At low frequencies, this arrangement behaves like a conventional transformer, but at high frequencies, where parasitic elements predominate, the device behaves like a transmission line. If the winding inductance and capacitance have been properly proportioned, this approach results in greater bandwidth than can be obtained from a conventional transformer design.

In the case of a two-winding transmission-line transformer operating between equal impedances, the characteristic impedance of the line, $Z_o = \sqrt{L/C}$, should be made equal to the terminating impedance. Under these conditions, the transformer would theoretically have infinite high-end bandwidth. However, the output hybrid of the basic-repeater power amplifier was required to operate between an amplifier

impedance of 50 ohms and a line impedance of 75 ohms with the feedback-port impedance equal to 150 ohms. Since the transmission line used to realize the hybrid could not simultaneously assume all these different values, the best compromise was to make Z_o equal to the geometric mean of the output impedances, or 106 ohms. The ratios of the terminating impedances also forced the use of a trifilar winding. The impedance properties of such windings were studied for various wire gauges, twist rates, insulation thicknesses, and materials. Ultimately, a transmission line made from three strands of AWG 40 polyurethane-insulated wire, combined at a rate of 35 twists per inch, was selected. Although other combinations would have produced the same Z_o, the high twist rate was used because it resulted in the best uniformity from model to model. The schematic diagram and transmission characteristics of the output hybrid are shown in Fig. 28. The divergence in insertion-loss characteristics at high frequencies is the result of the compromise value of impedance selection for Z_o.

Although transmission-line transformers were preferred from a component-design standpoint, they could not always be used because their feedback-port impedance must be an integral multiple of *both* the amplifier and line impedances. To overcome this constraint, conventional layer-wound hybrid transformers were used at the input
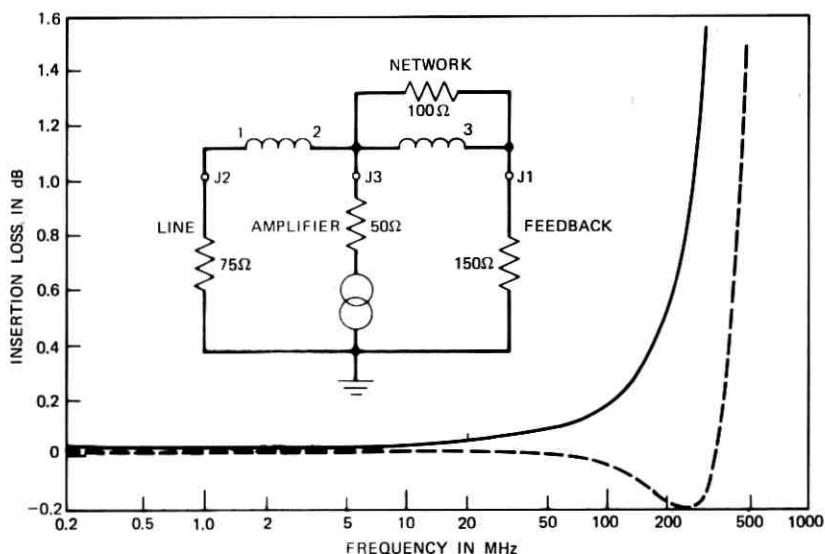

Fig. 28—Schematic and insertion-loss characteristic of output hybrid for basic repeater power amplifier.
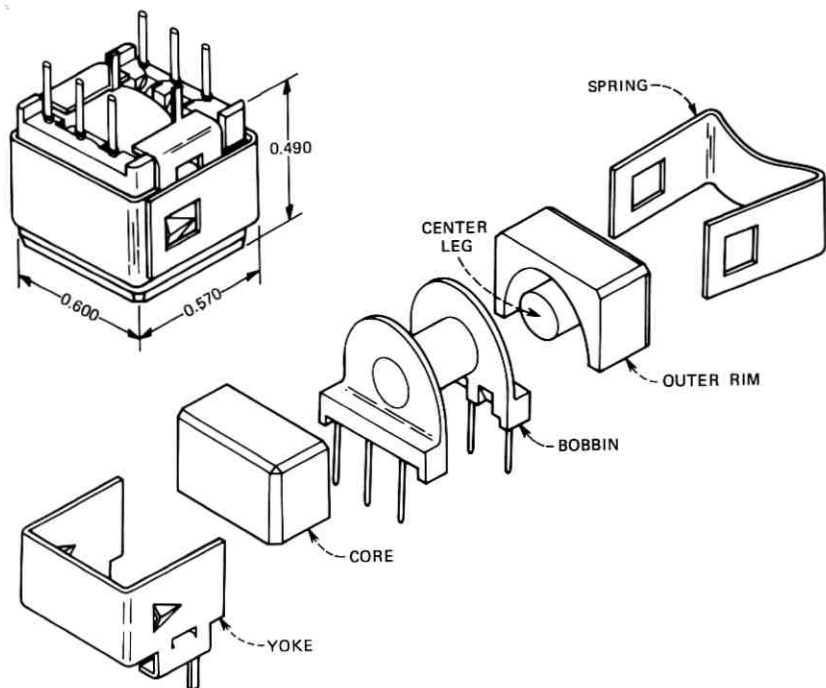
Fig. 29—Transformer structure for conventional layer-wound hybrids using D-cores.

and output of the basic-repeater preamplifier. The initial estimates of the impedances both transformers would be required to match were 75 ohms at the line port, 60 ohms at the amplifier port, and a "low" impedance at the feedback port. Various ratios were evaluated, and the final ratio selected was $75:65 + 28$ ohms, which results in a feedback-port impedance of 19.6 ohms.

To provide the best consistency in a layer-wound transformer, a structure that includes a winding bobbin is preferred. Bandwidth considerations, however, dictate that the structure would have to approach the efficiency of a toroid having a winding wrapped directly on the core and extending around its entire periphery. Core and winding dimensions, therefore, are nearly coincident. The D-core shown in Fig. 29 was chosen because the large area of the outer shell produces a low magnetic reluctance. This means that the center-leg length and area control the magnetic properties of the core. Since the winding bobbin fits directly over the center leg, the winding dimensions are only slightly larger than the effective core dimensions, and the structural efficiency approaches that of the toroid.

Evaluation of amplifier models revealed that their high-frequency gains were extremely sensitive to parasitics in the transformer windings. Requirements specified a ±0.02-dB reproducibility from model to model up to 70 MHz. Stringent winding procedures and in-process checks were required to obtain this control in a conventional transformer. Wire gauges and pitches were selected to provide smooth, even, single-layer windings. Carefully controlled paper insulation was used between the most critical windings to insure that the parasitic capacitance could be held to within ±5 percent, which is three times more stringent than normal manufacturing tolerances. Several different types of winding machines were evaluated to obtain one capable of providing the controlled-pitch and constant-tension features required to maintain performance tolerances.

2.3 Terminal components[1],[25]

One of the more difficult designs required for terminal applications was a 75:300-ohm, unbalanced-to-balanced transformer intended for use in the JMX modulators. The insertion loss of this transformer was to be held flat to within ±0.1 dB from 0.5 to 70 MHz, and the balance of the center-tapped winding was to exceed 50 dB from 0.5 to 90 MHz. Originally, separate designs were proposed for each jumbogroup because no core material was available that would permit simultaneous realization of these two requirements. In addition, an ability to maintain a balance of the required magnitude over this broad bandwidth had not been demonstrated at these high frequencies. After carefully studying the balance problem, however, it was felt that if an adequate core material could be developed and the mounting structure redesigned, it would be possible to cover the entire range with a single design by employing carefully positioned toroidal windings. The balanced winding consists of seven turns of a bonded pair of wires spaced evenly around the periphery of the core. These windings were then connected in a series-aiding fashion to complete the center-tapped 300-ohm winding. The 75-ohm winding was spaced between adjacent bonded turns and all leads carefully dressed, resulting in a sufficient degree of structural symmetry to maintain the balance. Simultaneously, a new core having the necessary properties was developed and the required mounting structure realized in a new physical design.

Hybrid transformers, to provide combining and splitting functions at a 75-ohm impedance level, were required for several different applications in the L5 system such as LPSS-3, the E3 equalizer, line-connecting arrangements, and JMX circuits. Mounting and coaxial-

connector variations resulted in four distinct physical designs, as shown in the photographs of Fig. 30. The basic electrical building block consists of a transmission-line hybrid and an autotransformer interconnected as shown in Fig. 30a. With reference to that schematic, autotransformer T1 matches the 75-ohm input at jack J1 to the 37.5-ohm impedance at the center tap of hybrid T2. It is wound with a twisted-pair wire whose characteristic impedance $Z_o = \sqrt{75 \times 37.5} = 53.0$ ohms to optimize performance. Since the impedances to be matched differ by $1:2$, the tap must be placed at a position corresponding to $1:\sqrt{2}$. This means that the windings cannot consist entirely of twisted-wire transmission line, but must include some free turns. These are created by partially decomposing the twisted bundle, with best results obtained by minimizing the number of free turns.

The hybrid T2 has a $2:1$ turns ratio and is wound with a twisted pair having a characteristic impedance of 75 ohms. The network terminating resistance R has a value slightly greater than the ideal 150 ohms to account for transformer core losses, and the capacitor C enhances the high-frequency response of the device. The electrical performance achieved by these components over the 1- to 70-MHz frequency range is as follows: (*i*) the transmission loss from input to either output is 3.10 ± 0.05 dB, (*ii*) the trans-hybrid loss between conjugate output ports is greater than 30 dB, and (*iii*) the return loss at any port is greater than 26 dB. In addition, the two output ports of any individual hybrid track to within 0.02-dB loss and 0.5-degree phase shift. An ideal hybrid would have 3.01-dB transmission loss, infinite trans-hybrid and return losses, and perfect flatness and tracking, with both outputs exactly in phase with each other and the input.

In addition to their use as individual hybrids, particular pairs of these transformers may be grouped to form large *n*-port arrays or "trees." The transformers in Fig. 30b use standard Bell System coaxial plugs and jacks so arranged that they may be interconnected directly. The transformers in Fig. 30c use miniature coaxial connectors and may be formed into trees by cabling them together. To reduce the number of interconnections required, the hybrid at the right of Fig. 30c, a dual hybrid, has an input, one -3-dB output, and two -6-dB outputs. The hybrid at the left of Fig. 30c is a single hybrid of the type shown in Fig. 30a.

2.4 Transmission measurements

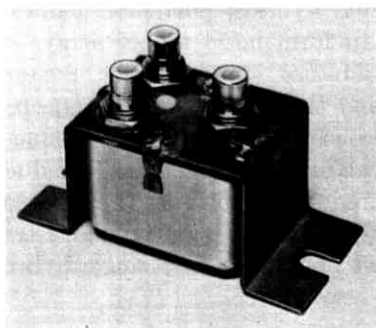The ± 0.02-dB reproducibility of insertion loss required for many L5 magnetic-component designs could not be guaranteed initially by direct measurements because neither the test set nor the apparatus
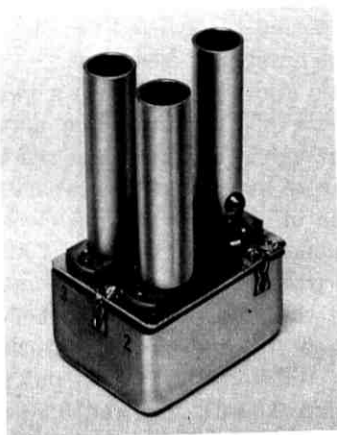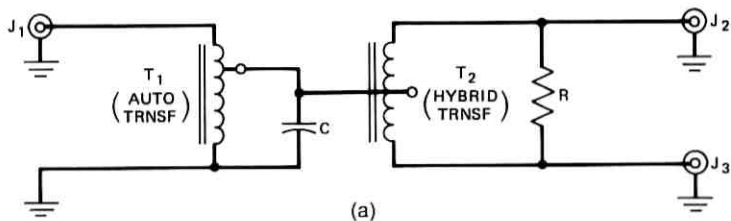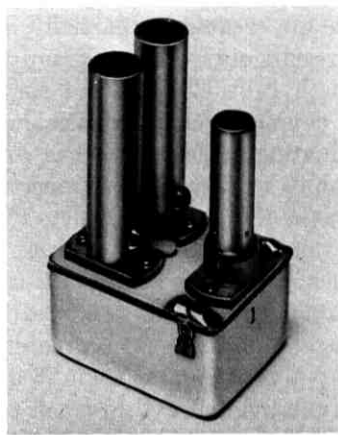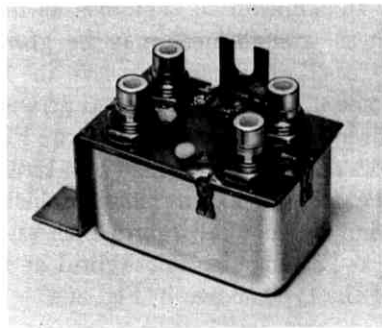
Fig. 30—Hybrid transformers for L5 terminals. (a) Building-block schematic. (b) Hybrids with Bell System coaxial connectors. (c) Hybrids with miniature coaxial connectors.

under test exhibited a return loss compatible with this extremely tight limit. Instead, a system was used in which a set of transformers known to have the proper characteristics in an amplifier was used to calibrate a test fixture. The calibrated fixture was then used to measure product. To guard against changes in the fixture or measurement system, however, these measurements were correlated to those made on reference transformers whose histories were well known. When either fixtures or reference units needed replacement, the appropriate steps in the calibration and correlation processes were repeated. Meanwhile, a program to provide improved measurement capability for these devices was begun.

To have a reproducible base line for loss measurements on COTMS,[3] a simple strap between the transformer input and output ports of the test fixture was used. While this introduces reflections because these ports often have different impedances, it is the simplest and most consistent method of eliminating differences in test sets, connectors, and cables. Furthermore, with sufficient padding close to the COTMS ports of the fixture, satisfactory base-line reproducibility can be obtained. Accordingly, alternate measurement schemes were discarded in favor of upgrading the existing approach.

The test fixtures were originally constructed with ordinary PWB's. The contacts to the pin-type transformer terminals were made with miniature spring sockets, and the interconnection to the test set used Bell System coaxial connectors. The resistive pads used to match the transformer impedances to the COTMS ports were made with various types of discrete resistors for the different transformer codes. It was recognized very early that the required reproducibility of measurements would be difficult to maintain in a production environment with this type of test fixture. Coaxial connectors can introduce impedance discontinuities, and spring sockets can become contaminated or wear out with frequent use. When replaced, resistive pads are often damaged by the heat of soldering and change value. In addition, the associated parasitic electrical element values are a function of exact mechanical configuration and placement which are difficult to control.

Improved designs of the spring contacts and of the coaxial connectors were introduced to improve the performance of the test fixtures. Knife-edge contacts with force-free insertion were used to replace the former, while the latter were superseded by precision 50-ohm versions. As the next step, the PWB was replaced by an alumina substrate having controlled dielectric properties. Tests quickly indicated new problems. Because of the higher dielectric constant of the substrate,
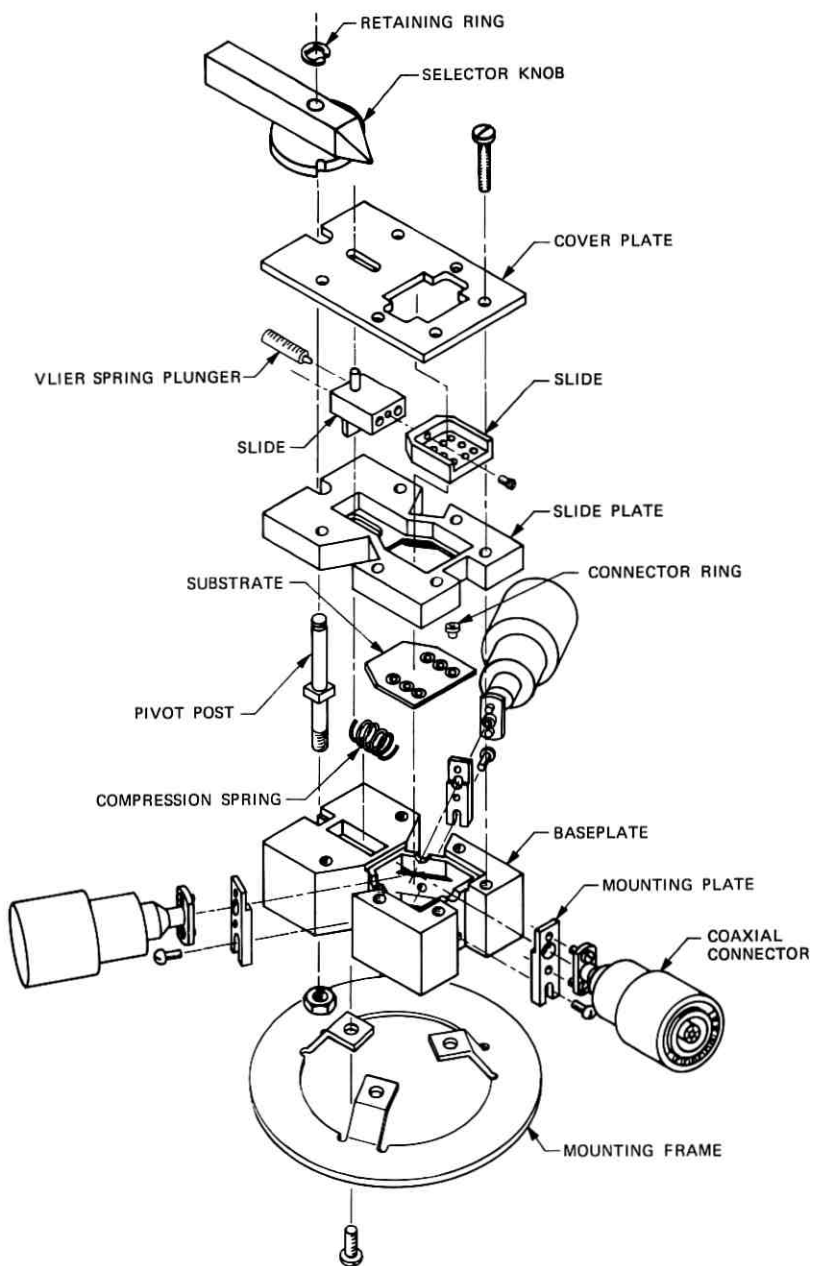
Fig. 31—L5 transformer test fixture.

the effective electrical length of the connections increased significantly, causing difficulty with crosstalk and reflections. The chip resistors, which were chosen for the matching pads, changed value on soldering and caused reproducibility and rework difficulties. A final design evolved in which the path lengths of the microstrip connections on the substrate were reduced to a minimum and the resistors realized as thin-film elements. This arrangement permits high accuracy in the adjustment of resistance values, makes parasitic elements closely reproducible, and eliminates the repair operations that had been required with the original fixture design. External connection problems were further reduced by coupling the coaxial connectors to the substrate via precision spring-loaded microstrip launchers. These new fixtures, shown in the exploded view of Fig. 31, have attained a reproducibility of better than ±0.01 dB.

III. CONCLUSION

Transmission networks and magnetic components are among the least visible components of the L5 system. Despite this lack of visibility, these components serve vital systems functions. Many of these functions are reviewed in this paper. The principal objectives of the paper, however, are to record those state-of-the-art advances that have contributed to the success of the L5 system, and to identify those allied technologies that have influenced network and magnetic component development. The influence of the computer in analysis and synthesis, precision measurement, and optimization is stressed, and the added influence of development in the piezoelectric device and magnetic materials areas has been noted. In many instances, developments in these allied technologies predated the L5 system and resulted from a general philosophy of "tool building" initiated, specifically, to have these technologies keep pace with future systems and component-development needs. Without such a philosophy, L5 requirements could not have been met on schedule.

IV. ACKNOWLEDGMENTS

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L5 SYSTEM:

Role of Computing and Precision Measurements

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The L5 Coaxial-Carrier Transmission System is the first long-haul, high-capacity transmission system for which the design was strongly influenced by extensive application of computer-aided design (CAD) techniques. A tight time schedule required a parallel effort of (i) improving and using the somewhat limited capabilities of existing CAD programs and (ii) developing new programs having increased capability and efficiency. Similar development of computer-controlled measurement techniques provided necessary device and component characterization and subsystem evaluation. The result is a powerful group of tools that are independently important, but whose combined use helped make possible the timely completion of the L5 system design. These tools, now "proved in," will profoundly influence the next system design philosophy.

I. INTRODUCTION

At the beginning of the L5 project in 1968, the set of computer-aided design (CAD) tools and the specialized computer-operated transmission measurement set (COTMS) at the Merrimack Valley location of Bell Laboratories were in an early stage of evolution. The tight system development schedule prevented the thorough, leisurely development of sophisticated programs that could provide the complete analysis and characterization needed for high confidence in design. Rather, a parallel effort was undertaken to augment existing design aids and to develop improved ones. The result was immediate answers generated inefficiently for early designs and a powerful group of programs that

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have been "proved in" by use during the later stages of L5 development. These were, however, not completed early enough in the design cycle to allow their full impact to be exerted on the overall design process in the manner that is now possible.

The following sections illustrate typical uses of computer programs for seven phases of the design sequence. These include:

(i) Small signal ac analysis of circuits having many nodes, including sensitivity and tolerance analyses.
(ii) Optimization of element values in a circuit.
(iii) Statistical measurements of early production models of repeaters for use in equalizer design.
(iv) Nonlinear distortion analysis.
(v) Component characterization.
(vi) Overall system analysis.
(vii) Equalizer characterization and alignment during manufacturing.

Limitations as well as successes are included to illustrate why and how the evolution of programs took place.

II. SMALL-SIGNAL AC CIRCUIT ANALYSIS

Several general-purpose circuit analysis programs,[1] as well as optimization[2] and Monte Carlo[3] programs, were in wide use in the communications industry by 1968. However, none was able to handle a circuit of the size (over 80 nodes) and complexity of the L5 repeater. While the next generation programs were being developed, modification of the existing programs and segmentation of the circuit were necessary to solve the immediate circuit analysis problems.

Since the L5 system uses amplitude modulation, the repeater must be a very linear device; this places strong emphasis on accurate, small-signal ac analysis of all circuits in the signal path. The basic repeater amplifier alone contains 80 nodes, 111 passive components, 8 transistors, and 4 wideband transformers. These numbers do not include "parasitic" elements found to be essential for accurate modeling so that, in practice, the actual node and component count must be significantly increased.

This section highlights some of the practical difficulties encountered with large circuits and briefly describes the next generation program that evolved from these experiences.

2.1 Partitioning large networks for analysis

To overcome the size problem, the circuit was at first partitioned into blocks that could be analyzed with existing programs. One output of each of these analyses was a multiport admittance matrix characterizing the block at those nodes where it interconnects with other circuit blocks. These block representations were stored in a library for subsequent retrieval when the performance of a multiblock circuit was to be calculated. This approach of using a frequency-dependent matrix to represent subcircuits proved useful, not only for reducing node and element count, but also for allowing the direct use of measured data from COTMS to characterize components not easily or accurately modeled by lumped element values.

The circuit segmentation approach also aided the calculation of sensitivities to element values. As each element in a given block was varied to determine overall circuit sensitivity to that element, the matrix representations of other blocks remained unchanged, saving much calculation time for the overall problem. Sensitivity data proved useful in many ways. Since a full-scale Monte Carlo analysis of the circuit was impossible during early development, a linear sensitivity model was used to set component tolerances based on their predicted effects on frequency response. The sensitivity data on parasitic element values also identified critical areas of the circuit where careful modeling and control were essential. It further identified those elements to be subjected to optimization and smaller-scale Monte Carlo analysis.

The segmentation approach also permitted use of the early optimization and Monte Carlo programs.[4] These were particularly helpful in the design of the feedback networks which, though small compared to the overall repeater, were critical in determining the frequency shaping of the repeater.

Although the "divide and conquer" method did allow some success in the early stages of analysis, several difficulties became apparent that were corrected in the next generation programs. The many steps involved in the piecemeal approach led to analysis turnaround times of the order of one day for a complete circuit. This severely limited the number and type of "what if" questions that could be raised and answered. Not only was the overall turnaround time relatively long, but all the segments had to be run in a batch mode, as no interactive facilities were available for the designer to pose such "what if" questions quickly or effectively. These inefficiencies, coupled with the non-static nature of the actual circuit as various engineers wanted to

implement changes suggested by earlier analysis steps, made the turnaround time a serious problem. The multiple passes through various analysis segments also led to interfacing problems and increased probability of coding errors.

2.2 A new ac analysis program

The program developed to overcome most previous limitations is called circuit analysis program for efficient computer-optimized design (CAPE-COD). It achieves its power and efficiency through a sparse matrix technique and special, machine-generated coding. It has analyzed circuits with 120 nodes, provides a wide variety of engineering-oriented outputs, allows simple problem formulation by engineers, and provides interactive capability on a graphics terminal.

The formulation technique chosen for CAPE-COD uses the tableau[5] approach. This method of arranging the system of equations that defines the topology and behavior of a circuit not only yields an efficient analysis scheme, but also gives flexibility and simplicity to the program design while minimizing the restrictions placed on the circuits that can be analyzed. For example, the following types of elements that are seldom allowed in ac analysis programs *are* allowed in CAPE-COD: controlled voltage and current sources; zero-valued resistors, capacitors, and inductors; and devices that have no Y or Z matrix representations, but have an S matrix representation (e.g., an ideal transformer or a three-port ideal circulator). By inserting zero-valued components into the original circuit where one suspects parasitics to be important, the program can then systematically investigate their effects without recoding, simply by varying the values from zero on subsequent runs.

The elimination order for the tableau can produce, at one extreme, essentially a nodal formulation; at the other extreme, a mesh formulation; in general, a hybrid formulation results. This flexibility gives the tableau method the capability of producing more accurate results than other methods. Such accuracy improvements are important when an element or subcircuit has a Y or Z matrix representation that is ill-conditioned, e.g., the Y matrix of a very small resistance.

To minimize core memory requirements of the computer and to allow analysis of large circuits having over 100 nodes, sparse matrix reduction techniques are used throughout the program. For efficiency in execution time, a loopless machine code program to solve the tableau system of equations is generated first for each problem. This efficiency is most important when the analysis package is used as a

subroutine in an iterative program such as optimization or tolerance analysis, since the time spent generating the machine code is insignificant compared to the time for repeated analyses. In addition to this machine code, dynamic memory storage allocation and memory paging schemes are used to increase efficiency further.

Input data for CAPE-COD can be entered in completely free format, which not only provides great convenience for the user, but also reduces the number of coding errors in large circuits. The input "language" was developed to be natural for an engineer, easily understood and remembered. The CAPE-COD currently accepts the following circuit elements: resistors, capacitors, inductors, controlled or independent voltage or current sources, two forms of transmission lines, two models for transistors (one of which is used for distortion analysis, described later), and N-port black-box devices. The latter can be characterized by directly measured Y, Z, or S parameter data, or by a user-supplied subroutine that calculates these parameters. In addition to having fixed values, elements can be optionally defined as a tabular function of frequency. This black-box capability was essential to the L5 repeater analysis, since no sufficiently accurate lumped element model existed to represent the wideband transformers used in the forward and feedback paths. Instead, two-, three-, and four-port matrix representations were formulated from frequency-dependent measured data

Perhaps the most important improvement over previously used programs which CAPE-COD provides is the wide variety and flexibility in its computed outputs. These include any branch voltage or current, any node voltage, or any of the following transmission related quantities: insertion gain or loss; voltage, current, or power gain or loss; return loss or reflection coefficient; load impedance or admittance; driving point impedance or admittance; $\mu\beta$ gain and $\mu\beta$ impedances; and Y, Z, or S parameter matrix values.

The $\mu\beta$ gain, i.e., the loop gain of a feedback amplifier, is especially useful for verifying stability in the amplifier. Since many of these outputs are complex numbers, they can be listed in any of the following forms: real and imaginary parts, magnitude and angle (in degrees or radians), dB$[= 20 \log_{10}(\text{magnitude})]$ and envelope delay (the derivative of the angle with respect to angular frequency).

Sensitivity studies are facilitated by another feature of CAPE-COD. After an analysis with nominal circuit values has been completed, it is possible to modify the values of any (one or several) element, subject to constraints if desired. A second analysis is then done, and the output quantities can be the *difference* in any normal output that results

from the modification of the circuit element values. This provides a direct measure of circuit sensitivity to any change, large or small, in any element values.

III. OPTIMIZATION

For a system as complex as L5, no exact design procedure exists at any level, either device, circuit, or system. Many iterations are required from an initial to a "final" design. Various tradeoffs are weighed, merit criteria are assigned, and design parameters are adjusted to yield a result that is optimum under given assumptions.

Various optimization programs, including SUPROX,[2] have been used at Bell Laboratories and have achieved good results. However, as with basic ac analysis programs, shortcomings led to the development of a new general-purpose optimization program named GPOP. Its capabilities are described briefly, and some key applications to L5 design are given as illustrations of its use.

3.1 Program capability

The GPOP is capable of minimizing a criterion function $F(\mathbf{x})$ by adjusting the set of n parameters $x_1, x_2, \cdots, x_n$ (or in vector notation $\mathbf{x}$) which may be bounded (above and/or below), unbounded, or fixed. A simplified program flow chart is shown in Fig. 1. Built into the program is a least-pth error criterion function that has the following form:

$$F(\mathbf{x}) = \sum_{j=1}^{l} \sum_{i=1}^{m} w_{ij}[y_{ij}(\mathbf{x}) - r_{ij}]^p,$$

where

$y_{ij}(\mathbf{x}) = y_j'(\mathbf{x}, a_i)$ is the jth response function value evaluated for the ith independent variable a_i. For example, y_1', y_2', and y_3' may be loss, delay, and input impedance of a given circuit, while $\mathbf{x}$ represents the circuit parameters and a_i is the ith frequency point.

r_{ij} is the requirement of the jth response function for the ith independent variable. If a *range* of requirements is requested instead of single value, and if $y_{ij}(\mathbf{x})$ falls outside the range, the r_{ij} in the formula is the value closer to $y_{ij}(\mathbf{x})$.

p is any positive even integer.

w_{ij} is the weighting constant of the jth response function at the ith requirement point.
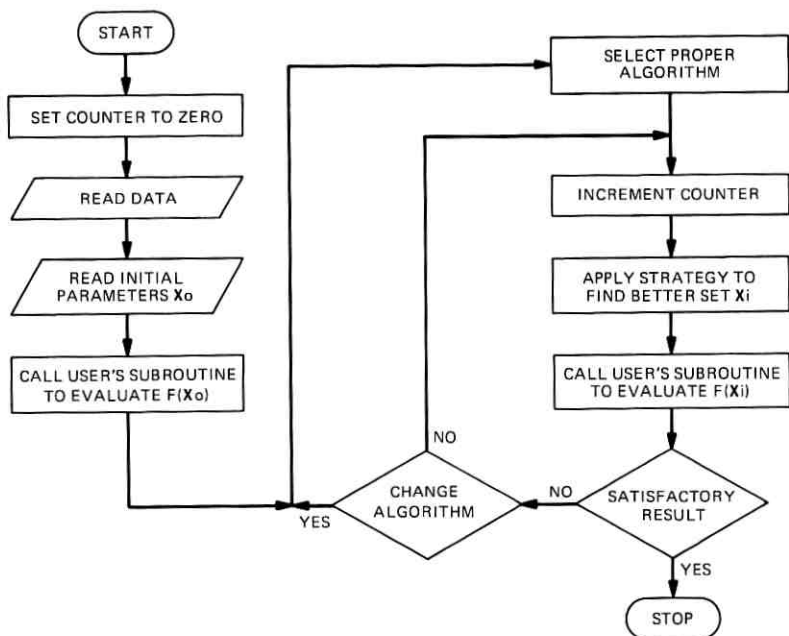
Fig. 1—Simplified flowchart for GPOP.

If the least-pth error criterion is not adequate for an application, the GPOP program also accepts a user-defined criterion function in the form of a subroutine. A user-supplied subroutine must be given to evaluate $y_{ij}(\mathbf{x})$ for the above error criterion. This subroutine is often in the form of a complete ac circuit analysis done by another program, such as CAPE-COD, but now treated as a subroutine.

A variety of strategies or algorithms exists for finding the optimum. Since the efficiency of these strategies is problem-dependent, it is desirable to have many available from which the user (or program itself) can choose. The GPOP is equipped with Fletcher-Powell,[6] steepest descent,[7] Nelder-Mead,[8] least-pth approximation[9] (least-squares when $p = 2$), and many others. Stopping the search or switching to another algorithm may be done either after a given number of iterations[*] or after the percent improvement is less than a specified value for five consecutive iterations.

Since the above algorithms work only for unconstrained optimization problems, the optimization of a bounded parameter problem (which is

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[*] In GPOP, an iteration corresponds to an improvement; i.e., in one iteration, the algorithm searches until a better criterion value is found.

a simple form of constraint) is accomplished by transforming the bounded parameters into free parameters. A variety of transformations have been built into the program. A user may also supply a specific transformation for a particular application. The gradient of the response function can be computed either from a numerical difference approximation or from a user-supplied expression. Input and/or output subroutines for pre- and/or postoptimization computation add another useful capability. In some problems, it is necessary to apply parameter constraints (such as interrelationships between parameters) independently of any parameter bounds. The input or output routines in GPOP allow the user to apply sequential unconstrained optimization techniques to constrained optimization problems.[10]

3.2 Application

The application of GPOP to a *system* design problem is given in a later section; here, a circuit problem, the optimization of a fixed deviation equalizer, is used for illustration. In practice, the fixed equalizer applies an average correction for the difference between repeater gain shape and cable loss shape.[11,12] Residual errors are then corrected in the field, using adjustable E1-E2 equalizers. The criterion function does not (and should not) consider the error at the output of the fixed deviation equalizer, but rather the final error after all
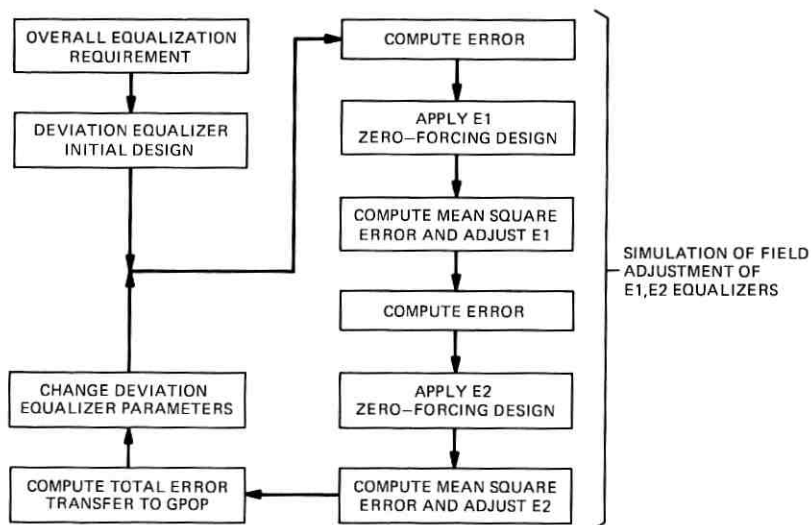

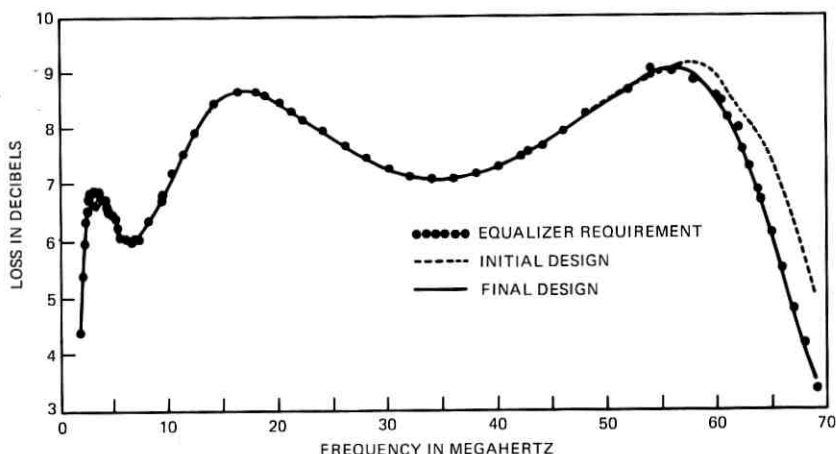Fig. 2—Flowchart for optimization of a fixed deviation equalizer.

Fig. 3—Performance of fixed deviation equalizer before and after optimization.

three equalizations have been applied. Thus, the program must simulate the field adjustment process before it can optimize the parameters of the fixed equalizer. Figure 2 shows the flow chart for this process, and Fig. 3 shows the result of optimization. Four stages of bridged-T equalizers, having 32 elements, were needed in the final design for the excellent match to requirements.

IV. STATISTICAL TRANSMISSION MEASUREMENTS OF REPEATERS FOR EQUALIZER DESIGN

The equalization hierarchy of the L5 system defined two levels of accuracy needed for transmission measurements on subsystems such as the basic repeater, regulating repeater, and equalizing repeater. These levels must be kept in mind to appreciate the measurement problems that had to be solved. First, the basic repeater provides a gain shape to compensate for cable loss (proportional to $\sqrt{f}$), going from 4.97 to 32.04 dB over the range from 1.6 to 66 MHz. The manufacturing goal was that basic repeater gains would not deviate from target shape by more than ±0.1 dB at the two-sigma points of the statistical distribution. Thus, a transmission measurement accuracy of about 0.01 dB at each frequency was needed. The wide dynamic range (30 dB) and frequency range (nearly two decades) made this 0.01-dB target difficult to achieve. Since measurements could be made on only a limited number of basic repeaters, but the mean shape had to be extrapolated to apply to many repeaters in the system, the accuracy of individual measurements should, if possible, exceed the 0.01-dB

target. Only a test set such as the COTMS[13] was able to provide this capability with the speed needed for a thorough job. Typical runs included 100 discrete frequencies across the band and took one minute per run.

The second level of measurement accuracy was needed for the remaining steps in the hierarchy, the regulating and equalizing repeaters. Since these provide mainly second- or third-order corrections, the dynamic ranges to be covered were relatively small, but accuracies needed approached 0.001 dB over the same wide frequency range. Nothing can be above suspicion at the 0.001-dB level (i.e., one part in 10^4). Mistermination or mismatch errors are the most frequent and insidious problem. Over this frequency range, even the best available pads needed to mask the test set impedances from those of the unknown showed significant frequency shaping of both transmission and impedance characteristics. Inexpensive mass-produced connectors, as used in field installations, have poor connection repeatability and impedance in terms of these accuracy requirements. But the temptation is strong for the infrequent user to assume that an instrumental resolution of 0.001 dB will guarantee him an absolute accuracy of 0.001 dB in any measurement. So regular accuracy verification of the test equipment, use of selected highest quality test fixtures and masking pads, and careful error analysis of each measurement was essential; when these checks were omitted, serious errors were often discovered later.

A third complication arose from the need for detailed temperature characterization of the repeaters. The maximum temperature coefficient of 0.0016 dB per °F for the basic repeater is a value critical in the design of the dynamic equalizer and the regulating repeater. A computer-controlled environmental test chamber was used with COTMS to accumulate sufficient data to determine this. Continuous control and automatic data logging minimized the operator interaction, but round-the-clock measurements required months to complete. Long interconnecting cables and a network of multiplexing relays introduced serious line reflections, requiring special care in selecting and locating masking pads. When the measurements were self-consistent, a standard statistical analysis program in another computer completed the temperature characterization.

V. DISTORTION ANALYSIS

Nonlinear distortion is a serious problem for a long-haul, frequency-division-multiplexed, analog communications system using solid-state

devices. Most serious are the second- and third-order intermodulation distortions, caused mainly by the transistors. Allowable distortions must be far below fundamental signals in practice; for example, the third-order distortion should not exceed a level 90 dB below fundamental. Calculation of such small distortions requires not only very accurate transistor modeling, but also appropriate mathematical tools to extract such a small fraction of the main signal.

5.1 A new distortion analysis program

Circuits having only a few transistors connected in cascade and a linear feedback path have been analyzed by S. Narayanan,[14] using the Volterra series approach. While answers are obtained in closed form, the computer implementation is not general and cannot handle the large complex circuits in an L5 repeater. So a new program was written to treat the second- and third-order distortions of transistor circuits under small signal conditions. It is called the nonlinear distortion analysis program (NODAP) and is based on the perturbation method.

For this type of problem, successful analysis depends on an accurate active device model. The integral charge control model (ICM) of Gummel and Poon,[15] which contains many nonlinear effects not included in conventional models, was chosen. Some more important effects include conductivity modulation, base push out, Early voltage, and avalanche effects. Expanding the ICM equations around a given bias point and expressing terminal currents in terms of junction voltages leads to an appropriate small-signal model. This is readily separated into linear and nonlinear parts, the linear part being representable by a hybrid-π equivalent circuit, or simply a two-port admittance matrix $Y(j\omega)$ (Fig. 4). Two controlled current sources, i_{Ne} and i_{Nc}, represent the nonlinear part. These together represent the intrinsic transistor, to which must be added appropriate parasitic elements; the parasitics are treated linearly except for the inactive collector capacitance.

The perturbation method works well (two iterations are sufficient), since the nonlinearities are small for small-signal conditions. The computational algorithm[16] uses any linear circuit analysis program (in this case, CAPE-COD). The result is an accurate, efficient, general-purpose program. The only nonlinear elements in the circuit are i_{Ne} and i_{Nc}, and they can be shown to be equivalent to two distortion current generators having amplitudes and phases determined by the second- and third-order nonlinear coefficients and the linear characteristics of the transistor circuit. This approach for a single transistor
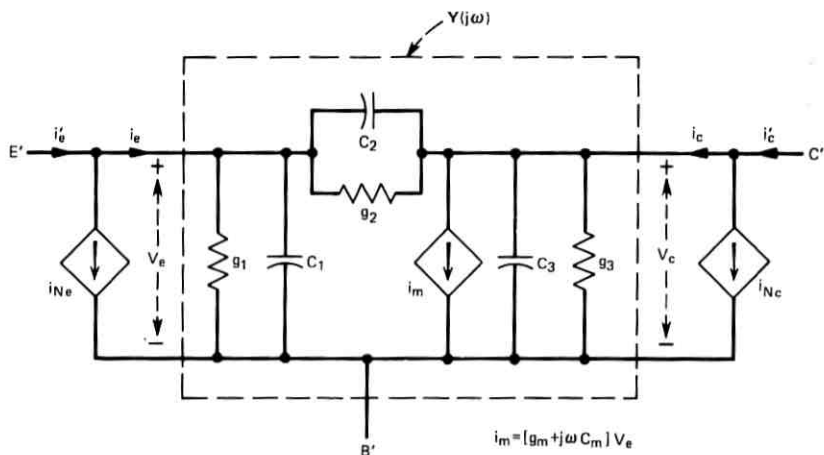
Fig. 4—Small-signal nonlinear transistor model.

can be readily extended to multitransistor circuits, since the distortion computation is carried out by a quasilinear approach and superposition holds.

The sequence of calculations in NODAP is interesting, showing how major programs can be combined. Given the circuit topology with the ICM parameters, bias point of each transistor, and sets of three fundamental frequencies (f_1, f_2, and f_3), NODAP computes the linear and non-linear coefficients for the transistor model. The linear characteristics are passed to CAPE-COD to calculate output voltages, load admittance, and junction voltages. These are returned to NODAP to calculate the second-order intermodulation current sources at $f_1 \pm f_2$, $f_1 \pm f_3$, and $f_2 \pm f_3$. These sources go back to CAPE-COD to calculate second-order distortion voltages. This type of looping is similarly repeated for the third-order distortion. The final step, by NODAP, prints out the calculated second- and third-order distortion indices M_{2E} and M_{3E}.

A "linear" and "nonlinear" option in NODAP allows the "turning off" or "turning on" of the distortion sources in any stage in an amplifier, to determine the contributions from that stage. The distortion effects of various nonlinearities can thus be isolated, giving insight impossible from direct measurements alone. Also, distortion *vectors* are computed, allowing the design of distortion cancellation circuits.

The calculated results for several L5 circuits have been in very good agreement with measured data, within 2 dB over a wide range of bias
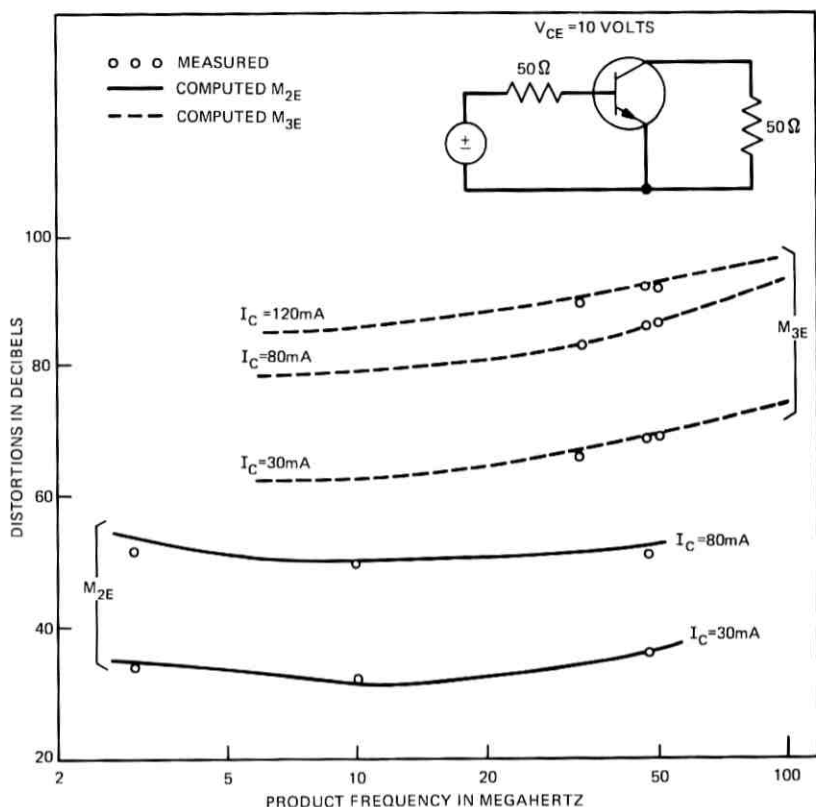
Fig. 5—Distortions of a common-emitter transistor amplifier.

conditions and frequencies. Figure 5 shows a comparison of predictions and measurements over the L5 frequency band.

5.2 Device modeling for distortion analysis

Accurate device modeling is essential to the success of NODAP in predicting nonlinear distortion. A program called extraction of parameters for the ICM of transistors (EPIC) has been developed to extract 30 ICM parameters from three sets of measured data. The raw data include the dc V-I characteristics, the junction capacitance characteristics, and the small-signal ac characteristics measured over a wide range of bias conditions. The previously described GPOP program is used to obtain the best ICM parameters by optimizing the match to the data. Those parameters associated with a single data set are matched first,

progressing to parameters that affect multiple data sets. This gives accurate results (physically realistic) quickly.

The main L5 transistor parameters were carefully extracted, using EPIC. In addition to their usefulness in predicting distortion, the linear model parameters were used in CAPE-COD for normal ac circuit analysis, giving excellent results. Agreement between computed and measured S parameters for the transistor was within 2 dB over the entire L5 frequency range.

VI. COMPONENT CHARACTERIZATION IN TERMS OF S OR Z PARAMETERS

Almost all the electronic subassemblies are built with components that are treated as classical "lumped elements" such as resistors, capacitors, inductors, or diodes, or as multiport devices such as transformers or transistors. Near the top L5 frequencies (66 MHz), the lumped element approximation is often inadequate and transmission line delays become important. Small capacitances and inductances, normally considered "parasitics," may have substantial effects in a circuit, especially over the extreme bandwidths covered by feedback amplifiers.

Simple passive components, such as two-terminal impedances, were measured automatically over frequency ranges up to 1000 MHz. Not only was the variation of element value with frequency obtained (caused by inadequate models), but derived properties were also obtained, such as the permeability and loss of magnetic materials, "Q" variation with frequency, self-resonant frequencies of reactive components, and linear two-element equivalent circuits of diodes. Such measurements were made quickly and very accurately on the COTMS, using an impedance calculating program. Impedances from milliohms to megohms can be measured over the range from audio to microwaves. Accuracies of the order of 0.1 percent are realized for real and imaginary components between 1 and 1000 ohms. A recent addition to the measurement tools is a computer-operated impedance bridge (COZY) that provides even greater resolution (approaching 10 ppm) and impedance range, though at frequencies limited to below 30 MHz.

The characterization of transformers and transistors as multiport devices was at a more sophisticated level. Basic measurements of voltage scattering (S) parameters were made automatically on COTMS, using a special program and calibration technique which compensates for the mismatch errors caused by imperfect test set impedances.[17] These corrections are especially important since test set impedances do not remain constant over wide frequency ranges. From the basic S

parameters, other matrix parameters such as H or Y were obtained by computer data transformation. The H parameters are particularly useful for transistors, while the Y parameters fit naturally into circuit analysis programs based on nodal techniques. The multiport transformers used in the L5 amplifiers were characterized by merging several sets of two-port measurements into a 4-by-4 Y matrix representation.

Many of these characterization jobs were not carried out until inexplicable problems appeared during the development of circuits. However, with the presently available tools that are quick, accurate, and easy to use, all critical component types can be characterized automatically before use in model circuits. Such routine checks can prevent many problems and save considerable time in the long run. A component as simple as a resistor has many nanohenries of body and lead inductance that can play havoc in high-frequency circuits, yet it takes only a few minutes to characterize the component with modern tools.

VII. SYSTEM DESIGN

In repeatered analog transmission system design, the optimal allocation of total system noise caused by various sources and subject to certain constraints is an important but tedious task, if the conventional approach is used. Manual trial-and-error modification of transmission levels and calculation of the resulting noise at various system points are time-consuming processes. All these are interdependent functions of repeater gain, noise figure, modulation coefficients, etc. A nonoptimum assignment may lead to failure to meet system design objectives and to possible redesign of basic elements, starting a new system design cycle. Since the process requires optimizing a performance function subject to a given set of system constraints, modern optimization techniques are applicable.

7.1 Computation of intermodulation noise

An intermodulation noise computation program called NEWMOD was written at Bell Laboratories in the 1960's for the development of submarine cable systems. This program employs a modified version of Bennett's product count method[18] in a formulation that takes into account the effect of the nonlinear phase characteristic of repeaters. The NEWMOD program was used extensively in the development of the L5 system, in which the introduction of phase shaping networks effectively randomized the addition of third-order intermodulation prod-

ucts. This allowed the channel capacity of the system to be increased by about 20 percent for the same noise performance.

The product count method requires too much computer storage and processing time to allow the coupling of NEWMOD with GPOP for system noise optimization. This difficulty can be avoided by modifying the earlier formulation so that a high-order Chebyshev numerical-integration formula is employed to replace the product count approach. Based on this new formulation, a program named noise optimization for repeatered analog transmission systems (NORATS) has been implemented and is briefly described.

7.2 NORATS program objective and organization

The main objective of this optimization is to select the best repeater output transmission level as a function of frequency for the given conditions. For a noise-limited system such as L5, the total system noise should be minimized. For an overload limited system, the average power output of repeaters should be minimized so that the total system noise still satisfies the requirements. In both problems, we may consider the modulation coefficients as additional design parameters subject to certain limitations. All these problems can be considered special cases of minimizing the following criterion function

$$F(\mathbf{x}) = \sum_{i=1}^{m} w_i[y_i(\mathbf{x}) - r_i]^p,$$

where

$y_i(\mathbf{x})$ = total system noise (in dBrnc0) at frequency f_i $i = 1$, $2, \cdots, m - 3$,

$y_{m-2}(\mathbf{x})$ = average value of the second-order modulation coefficient (in dB),

$y_{m-1}(\mathbf{x})$ = average value of the third-order modulation coefficient (in dB),

$y_m(\mathbf{x})$ = equivalent rms power of a single tone at repeater output (in dBm),

r_i and w_i are the corresponding requirements and weighting constants, p is a positive even integer, and the components of $\mathbf{x}$ are the coefficients of the polynomial representations of the transmission level and modulation coefficients as functions of frequency. The total system noise function in the equation is the power addition of thermal, second-order,

and third-order intermodulation noise. This criterion function is in a form ready to be used in the GPOP program described in Section II.

7.3 NORATS applications

Solutions to various design problems can be carried out using NORATS by choosing appropriate weighting constants in the criterion function. For a noise-limited system, the weighting constant w_m corresponding to the output power is set to zero, since this allows the optimum transmission level to be determined by minimization of total system noise, regardless of output power. A relatively large w_m corresponds to an overload-limited system. The weighting constants w_{m-1} and/or w_{m-2} are set to zero if the corresponding modulation coefficients are given instead of being treated as design parameters. Recent NORATS runs have confirmed that earlier designs approached an optimum.

VIII. EQUALIZER CHARACTERIZATION AND ALIGNMENT

The strategy of using distributed corrections for the separate temperature variations of the cable and the repeaters (regulation), and for deviations from nominal frequency shaping caused by manufacturing tolerances (equalization), is a complex one. The strategy depends on accurate characterization of the various equalizer and regulator circuits to be effective, and this was considerably simplified by the use of COTMS. One special program for alignment during manufacture deserves description. It illustrates that, in some problems, even the most sophisticated measurement equipment alone is not enough to fill the need. By incorporating within the computer-controlled test set an algorithm for adjusting a set of nonorthogonal controls, considerable time and cost savings are achieved.

The 4211A network generates a $\sqrt{f}$ shape for use in regulating the L5 system against changes in cable loss with temperature.[12] Six adjustable elements, which are not independent, are needed to match the resulting shape to requirements. While their individual effects are readily measured, the interdependence makes final adjustment during manufacture nearly impossible by trial-and-error methods. The problem could be solved on a separate computer, but this would require voluminous data transfer and lost time. The logical time and place to carry out the adjustment were simultaneously with the measurement process. The details of the procedure and its effectiveness are described in another article in this issue.[12] This particular combination of precision measurement, data reduction, computer prediction, and output

"instructions" to an operator, all done in a single machine, will be the way of the future for such complex tasks.

IX. CONCLUSIONS

The complex design problems to be solved in the development of the L5 system led to the writing of a series of new computer programs that are highly interrelated and yet very useful independently. CAPE-COD provides ac circuit analysis with flexible input and output, speed of computation, and the ability to handle large circuits with very few limitations of types of elements. Continued development is under way to extend its range of application into the microwave region and to improve its interactive and statistical capabilities. The GPOP program implements modern optimization techniques through a variety of algorithms and considerable flexibility in types of constraints allowed. Both GPOP and CAPE-COD contain links to simplify their combined use, making an extremely flexible combination. The NODAP introduces a new method of handling nonlinear distortion problems, working together with CAPE-COD. Input data for NODAP are provided by EPIC, a transistor model parameter extraction program that works together with GPOP. A new system design program, NORATS, also works with GPOP to optimize the noise performance of repeatered analog transmission systems.

The usefulness of computer-operated transmission measurement equipment was enhanced by the development of advanced impedance measurement and modeling programs. Automatic measurement of many repeaters under computer-controlled environmental conditions was facilitated, along with data links to transfer measurement and characterization data directly to another computer where they could be used in CAPE-COD. A specialized alignment program was written for use in COTMS.

While each of these new tools is important in its own right, their combination and coordinated development provided an extremely powerful set of tools. Clearly, the L5 project could not have been completed on schedule without these computer aids to design and measurement.

X. ACKNOWLEDGMENTS

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Y. L. Kuo, M.S. (E.E.), 1961, Oklahoma State University; Ph.D. (E.E.), 1966, University of California; Assistant Professor, 1966–1970, Purdue University; Bell Laboratories, 1970—. Mr. Kuo's primary interest is in the area of active device modeling and computer-aided analysis of nonlinear networks.

M. L. Liou, B.S., 1956, National Taiwan University; M.S., 1961, Drexel Institute of Technology; Ph.D., 1964, Stanford University;

Bell Laboratories, 1963—. Mr. Liou is presently supervisor of the Analysis and Interactive Computing Group providing analytical and computational support to the transmission system development at Merrimack Valley. His fields of interest have included system theory, numerical analysis, optimization, and computer-aided design of circuits and various components in radio and cable transmission systems. Member, IEEE, Eta Kappa Nu, Sigma Xi.

Michael M. Luniewicz, B.S.E.E., 1958, University of Massachusetts; M.S. (engineering), 1961, Northeastern University; Bell Laboratories, 1960—. Mr. Luniewicz has been engaged in circuit design and development for multiplex terminals and coaxial lines.

Robert E. Maurer, B.S.E.E., 1962, M.S.E.E., 1964, Ph.D., 1968, Northeastern University; Bell Laboratories, 1962—. In addition to supervising a group responsible for the development of analog multiplex equipment, Mr. Maurer has worked on the design of the equalizing repeater for the L4 system, the analysis and modeling of intermodulation distortion, the equalization of random channels, and analysis and exploratory development related to the transmission of high-speed digital signals over sharply band-limited analog channels. He presently supervises a system-planning group working on a new baseband digital system for exchange area application. Associate Editor, *IEEE Transactions on Communications.* Member, Tau Beta Pi, Eta Kappa Nu, Phi Kappa Phi, Sigma Xi.

Samuel Mottel, B.S.M.E., 1950, City College of New York; M.S.M.E., 1968, Newark College of Engineering; Bell Laboratories, 1952—. Mr. Mottel has been concerned with physical design of power equipment. He has worked on power for carrier systems, microwave systems, submarine cables, key telephones, ringing and tone plants. He supervises a group responsible for physical design of a variety of power equipment.

Joseph M. Nacci, B.S. (Physics), 1956, University of Rhode Island; Bell Laboratories, 1956—. Mr. Nacci has been active in the design and development of a wide variety of silicon transistors and integrated circuits. These include pnp transistors of all types, npn ultralinear transistors, silicon based capacitors, and pnp integrated circuits.

Sundaram Narayanan, B. Tech., 1960, Indian Institute of Technology, Kharagpur, India; M.S., 1963, and Ph.D. (Electrical Engineering), 1965, Carnegie-Mellon University; Bell Laboratories, 1965—. Mr. Narayanan has worked on nonlinear distortion in transistor amplifiers and the use of high-speed digital signals over band-limited analog channels. He was supervisor of a group developing precision signal source and is presently supervisor of a group working on a new multiplex arrangement for L5 and a basic repeater design for an advanced coaxial system. Member, IEEE, Sigma Xi.

James F. Oberst, B.E.E., 1964, Manhattan College; M.S., 1966, and Ph.D. (Electrical Engineering), 1969, Polytechnic Institute of Brooklyn; Assistant Professor of Electrical Engineering, Polytechnic Institute of Brooklyn, 1968–1969; Bell Laboratories, 1969—. Since joining Bell Laboratories, Mr. Oberst has worked on various aspects of PCM transmission over cable and synchronization for FDM terminal equipment. He is presently working on PCM channel banks.

Arthur Olsen, Jr., B.S.E.E., 1959, Worcester Polytechnic Institute; M.S.E.E., 1961, Northeastern University; Bell Laboratories, 1959—. Mr. Olsen has been responsible for the design and development of transmission networks and magnetic components. He presently supervises a magnetic components group.

Edward J. Panner, B.S.E.E., 1962, Lafayette College; Bell Laboratories, 1949—. Mr. Panner has been engaged in device development principally for transmission systems involving technologies ranging from klystrons and general-purpose tubes to transistors and integrated circuits. Member, Tau Beta Pi, Eta Kappa Nu, Phi Beta Kappa.

Henry S. Pustarfi, A.E., 1955, Newark College of Engineering; Bell Laboratories, 1951—. Mr. Pustarfi has worked on the development of quartz crystal filters, temperature-control circuits, and thermoelectric ovens. He is presently engaged in the development of crystal-controlled oscillators, temperature-control devices, and precision frequency standards.

Richard W. Sanders, B.S.E.E., 1959, University of Vermont; M.S.E.E., 1961, Northeastern University; Bell Laboratories, 1959–1972; Western Electric Company, 1972—. At Bell Laboratories, Mr.

Sanders was engaged in the development of coaxial transmission systems, specifically protection switching for the L4 and L5 Coaxial-Carrier Transmission Systems. At Western Electric, he is presently involved in test engineering for the voice-band interface frame and the digroup terminal for the No. 4 ESS project. Member, IEEE, Tau Beta Pi.

T. H. Simmonds, Jr., B.S.E.E., 1954, University of Virginia; M.S.E.E., 1961, Northeastern University; Active Naval Reserve, 1954–1958; Bell Laboratories, 1954 and 1958—. Mr. Simmonds' early work was on a variety of filters and networks for long- and short-haul carrier transmission systems. As supervisor of a networks group in the Transmission Systems Networks Department he is responsible for work on transmission filters and networks for carrier and radio transmission systems. Member, IEEE, Tau Beta Pi.

Robert P. Snicer, B.S.E.E., 1966, and M.S.E.E., 1967, Massachusetts Institute of Technology; Bell Laboratories, 1967—. Since joining Bell Laboratories, Mr. Snicer has been engaged in transmission system development and computer-aided design. He presently supervises a network design group. Member, IEEE, Eta Kappa Nu, Tau Beta Pi, Sigma Xi.

John L. Thomas, B.S.E.E., 1957, University of Maine; M.E.E., 1960, New York University; Bell Laboratories, 1957—. Mr. Thomas engaged initially in circuit design work associated with special applications of submarine cable systems. He worked on systems analysis and supervised a group responsible for the circuit design of shore terminal transmission facilities associated with the SF submarine cable system. He later supervised a group responsible for repeater, equalizer, and special test set circuit design for submarine cable systems. He is presently responsible for the design of transmission surveillance and fault location circuitry for the L5 coaxial system. Member, Phi Kappa Phi, Tau Beta Pi.

Edward D. Walsh, B.S.E.E., 1965, Gannon College; M. Eng., 1966, and Ph.D., 1968, Rensselaer Polytechnic Institute; Bell Laboratories, 1968—. Mr. Walsh engaged initially in the frequency domain characterization of high-frequency active and passive devices. He has

developed a general-purpose frequency domain simulation program for transmission circuits. He is presently working on Monte Carlo simulation programs for transmission circuits. Member, IEEE.

R. J. Wirtz, B.S. (M.E.), 1950, Brown University; Bell Laboratories, 1956—. Mr. Wirtz was at first involved in resistor development and the physical design of the N3 Carrier System. He later supervised the physical design of the L4 Coaxial System. He is currently supervisor of the Coaxial System Physical Design Group responsible for the design of long-haul carrier systems.

Donald J. Zorn, A.A. (electronic engineering), Wentworth Institute, 1959; B.S. (industrial technology), Northeastern University, 1965; Bell Laboratories, 1959—. Mr. Zorn has worked in the development of A5 channel banks, L-multiplex, MMX-multiplex, and the L4 coaxial system, and is currently working on the L5 coaxial system. In addition, he has worked on several special development projects, including a carrier supply for channel banks used in the initial *Telstar* operations and a special pilot supply for the NORAD headquarters carrier system. Member, Sigma Epsilon Rho.