The L3 Coaxial System

Television Terminals

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Television terminals are required at circuit ends of the L3 coaxial system; at the transmitting end to condition video signals for carrier transmission and at the receiving end to detect the transmitted signals. Special signal characteristics, e.g., a degree of modulation which exceeds the value commonly referred to as 100 per cent modulation, require departures from standard modulating and detecting processes. The high degree of modulation requires both careful control of transmitted wave form and at the receiver product demodulation with phase synchronous carrier (homodyne detection).

Carrier regeneration requirements result in the choice of one step frequency translation from the video frequency spectrum to the allocated vestigial sideband carrier spectrum. The one step process using a single modulator results in unusual balance requirements for the modulator itself and an unusual circuit configuration.

Transmission quality objectives for the terminals are such as to permit six pairs of television terminals to operate in tandem in a transcontinental circuit. This permits a degree of interconnecting flexibility in operation with other systems, e.g., L1 coaxial or microwave systems. These objectives place severe requirements upon the transmission stability of various filters and other circuits within the terminals. New network techniques both in design and fabrication are brought to bear in the effort to achieve required performance.

The transmitting and receiving terminals are described, illustrating the functional operation and mechanical and electrical arrangements of the equipment.

Introduction

The main features of the L3 coaxial cable transmission system have been described in a companion paper.[1] This paper describes the television transmitting and receiving terminal equipment of the L3 system. The transmitting terminal conditions a television signal for carrier transmission over the system simultaneously with a group of 600 telephone messages. The receiving terminal reconverts the carrier signal at each receiving point along the cable route. Primarily, the transmitting terminal is a modulator which translates the composite video picture spectrum of frequencies up to the carrier band of frequencies and the receiving terminal is a detector which retranslates the carrier spectrum back to its original band of frequencies.

Particular characteristics of the transmitted television signal, which are intended to aid in achieving optimum transmission quality, have necessitated the departures from past techniques in modulation and demodulation processes that are described in the following. Described also are the methods employed to achieve transmitted picture quality adequate for tandem operation of as many as six pairs of transmitting and receiving terminal equipments in a 4,000-mile television transmission circuit. Operation with several pairs of terminals in tandem occurs when L3 coaxial systems are interconnected with L1 coaxial systems or microwave radio systems.

L3 television terminal development has been in progress since early in 1948. Two transmitting and two receiving terminals have been built on a preproduction basis and currently are being tested under field conditions as part of the L3 system field trial. Development effort is continuing on the terminals with emphasis on equipment reliability, including means for maintaining and improving transmission quality.

Frequency Allocations

The L3 coaxial system was designed to have as broad a transmission band as the economics of repeater spacing together with presently realizable feedback amplifier performance permit.[2] The band extends from 300 kc to 8.5 mc. In comparison with this band a broadcast television signal occupies the frequency spectrum from zero frequency up to 4.5 mc.

From the foregoing it is evident that the television spectrum will not occupy fully the available system transmission band. It is feasible and attractive to allocate part of the transmission band for television transmission and the remainder for transmission of message channels. Detailed allocations then result from a compromise among transmission performance, cost and the number of message channels made available.

From these considerations vestigial sideband transmission of the television signal rather than double sideband transmission is called for. The smaller the vestigial band of transmitted frequencies is made the smaller will be the total television band required. However, both the cost of band shaping filters and the difficulty of maintaining satisfactorily low values of vestigial sideband quadrature distortion increase as the vestigial band width is reduced. The compromise of these factors resulted in the choice of a 500-kc vestigial sideband.

Another choice made was to transmit the television signal in the upper part of the L3 band and the message channels in the lower part. This allocation was determined by considering the noise distribution in the transmission band together with the modulation distortion, (harmonic distortion), produced by the repeaters. By transmitting television in the upper part of the band a minimum modulation distortion is achieved since the harmonics of the television signal largely fall outside the transmitted band or at high frequencies where their effects in the picture are relatively less visible than low frequency distortion. This factor outweighs the higher noise level in the upper part of the band.

With respect to the television carrier location, it is placed at the bottom of the television band at 4.139 mc, with the vestigial sideband extending down to 3.64 mc and the main sideband extending upward to 8.50 mc. Alternatively the carrier could have been located at the top of the L3 band with a main lower sideband and vestigial upper sideband, but this choice would be disadvantageous because of higher noise levels and poorer repeater gain stability near the top edge of the transmitted band.

The final allocation of frequencies is shown in Fig. 1. Just below the television band from about 3 mc to 3.5 mc is a dead space. This frequency space is needed for the filters which are employed to separate the telephone signals from the television signals at television and telephone dropping points. A very high value of loss is required of these filters in their attenuating bands and if this is built up over too small a frequency

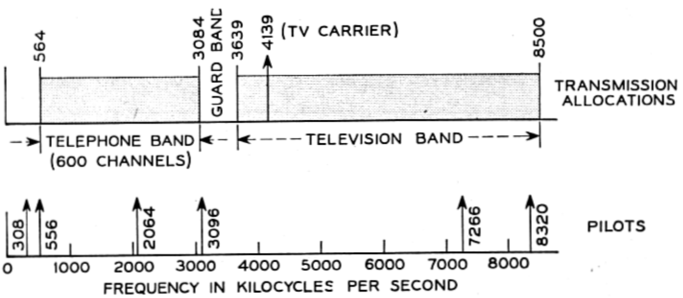


Fig. 1 — Frequency allocations for L3 combined television-telephone transmission.

band the resulting delay distortion introduced into the television band becomes very difficult and expensive to equalize. Below this "guard band" is the "master group" of 600 telephone channels which is transmitted simultaneously with the television signal.

The detailed allocation of specific frequencies, e.g., TV carrier and the pilot frequencies, results from consideration of effects produced by these frequencies in telephone channels as a consequence of modulation distortion in repeaters. These considerations are described in detail in the system design paper.[1]

Modulation Process

In the L1 coaxial cable system[3] frequency translation by the television terminals is accomplished in two stages. The two step process employs a first modulator supplied with a very high frequency carrier to translate all video frequencies to a band far outside the final transmitted band of

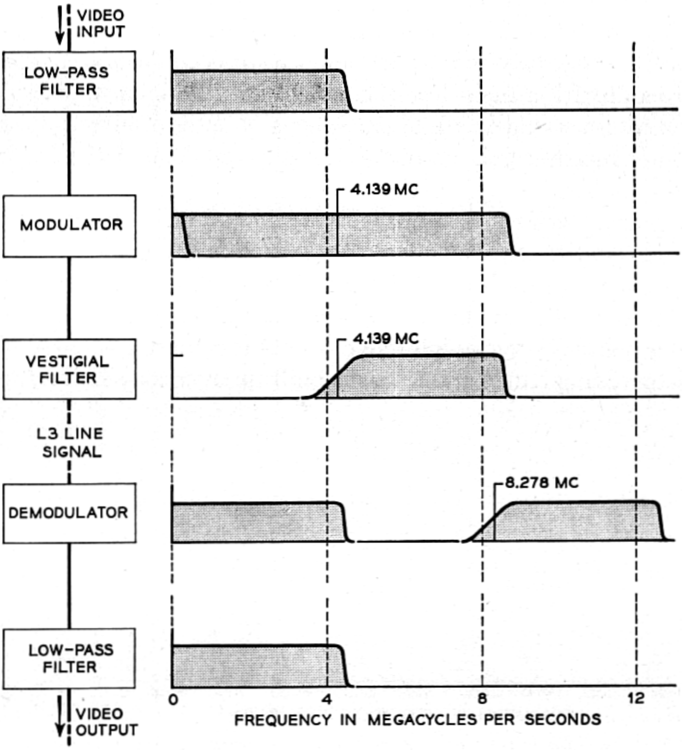


Fig. 2 — Television terminal modulation processes.

frequencies where the upper side band is suppressed. A second modulation then translates the vestigial sideband signal back down in frequency to the final band. In contrast the L3 terminals employ only a single step of modulation to convert the signal directly to the assigned band as shown in Fig. 2. In general this can be accomplished if the carrier frequency is at least half the sum of input and vestigial bandwidths. Then the lower modulation sideband does not fold over the zero frequency axis to produce frequencies which fall back into the vestigial or upper sidebands. Some foldover is evident in the L3 case shown in Fig. 2 at low frequencies of the modulator output. Single step modulation is advantageous in that the very high frequencies encountered in the multi-step process are avoided. The disadvantage of the one step process is that many extraneous products of modulation, which in the two step process can be suppressed with filters, must be reduced to tolerable levels by balances in the modulator.

The modulator, Fig. 3, is a combination of two double balanced modulators of a form often employed for modulation of telephone signals.[4] The effect of a double balanced varistor modulator of the type represented by either of the two in Fig. 3 is to multiply the input signal by a square wave function having the period of the particular carrier frequency. Since the square wave contains all odd harmonic multiples of

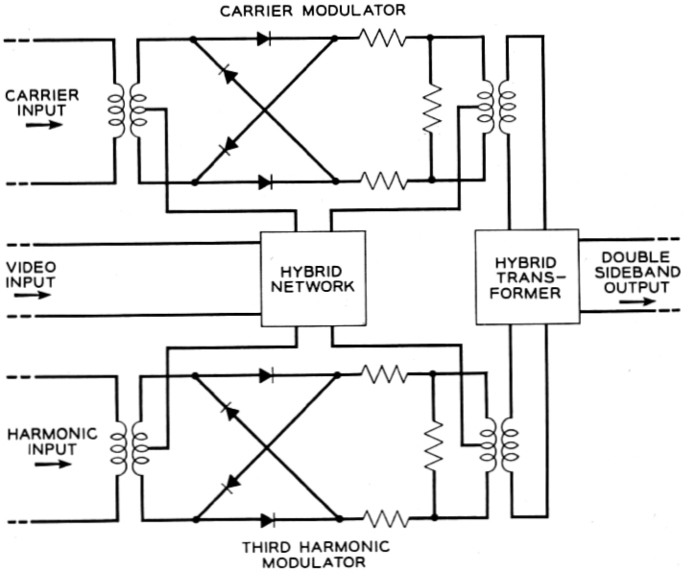


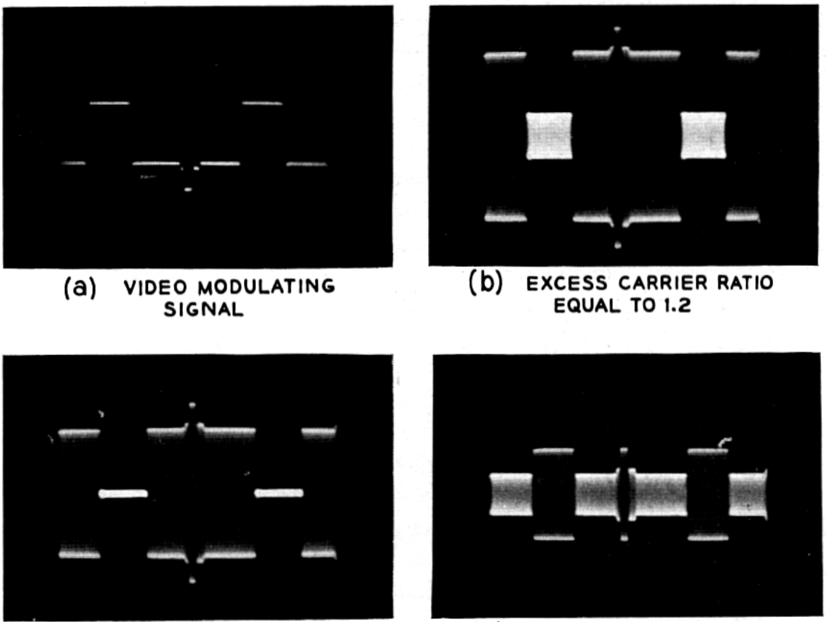
Fig. 3 — Modulator for L3 terminals with third harmonic carrier product balance feature.

the carrier frequency its multiplication by the input signal generates a series of double sideband output spectra, each centered about one of the harmonics of the carrier. It happens in this case that the lower sideband of the third harmonic spectrum of the carrier modulator contains frequencies low enough to overlap the high frequencies of the carrier spectrum upper sideband and this overlap results in quite visible picture distortion.

It is possible, by employing a second modulator paralleling the first but driven with a frequency three times the carrier frequency, to generate a signal spectrum centered at carrier third harmonic which will cancel the corresponding output of the carrier modulator. Successful translation of the video spectrum to the L3 carrier band in a single modulation step depends upon the maintenance of this and other modulator balances to unusually stringent requirements.

The carrier supply oscillators for the modulators and demodulators are of the Meacham bridge type[5] with quartz crystal frequency control and thermistor amplitude control. Frequency stability of two parts per million is required for successful carrier regeneration at the receiver. A constant temperature oven for the quartz plus the inherent stability of the bridge type circuit is expected to provide the required frequency stability between monthly maintenance periods.

A feature of the signal transmitted over the L3 system is a degree of modulation which exceeds the value commonly referred to as 100 per cent modulation. The resulting waveform contains a maximum ratio of information to peak carrier, important from the standpoint of optimum signal to noise performance. Fig. 4 shows progressively the reduction in peak carrier amplitude which may be effected by subtraction of carrier component from a modulated signal. Figs. 4(b), (c) and (d) each contain the same amplitude of video modulation. Fig. 4(b) represents a video modulated carrier signal with maximum carrier occurring at tips of synchronizing pulses and a minimum carrier, equal to 20 per cent of maximum carrier, corresponding to picture white. Fig. 4(c) represents the same signal as Fig. 4(b) except that the 20 per cent excess carrier has been subtracted. This is the 100 per cent modulation case. Fig. 4(d) shows the effect of further carrier subtraction, (addition of negative carrier), to reduce to a minimum the peak amplitude of the modulated signal. The waveform of Fig. 4(d) employed for L3 transmission requires 7½ db less maximum carrier power than that of Fig. 4(b) for the same transmitted information. The term "excess carrier ratio" has been devised to describe degrees of modulation which exceed 100 per cent. It is the ratio of peak carrier amplitude to the peak-to-peak modulation



(a) VIDEO MODULATING SIGNAL

(b) EXCESS CARRIER RATIO EQUAL TO 1.2

(c) EXCESS CARRIER RATIO EQUAL TO 1.0

(d) EXCESS CARRIER RATIO EQUAL TO 0.5

Fig. 4 — Carrier waves variously modulated by a composite video signal.

amplitude. Excess carrier ratio, (ECR), for the waveforms of Figs. 4(b), (c) and (d), respectively, are 1.2, 1.0 and 0.5.

Modulated signals of the forms of Fig. 4(b) or 4(c) may be detected by rectification, i.e., envelope detection. However, rectification of the waveform, Fig. 4(d), produces a spurious envelope wherein video signals which exceed a particular value are inverted. It is necessary to employ homodyne detection, that is, a demodulator driven by a locally generated carrier which is synchronous in phase angle and frequency with the carrier component of the signal wave. As described later, homodyne detection also makes possible the necessary suppression of the quadrature distortion associated with vestigial sideband transmission. Quadrature distortion associated with envelope detection is tolerated in the L1 coaxial system but with tandem operation of terminals required in the L3 system, would accumulate to intolerable values.

Vestigial Sideband Considerations

A vestigial sideband signal is produced by a band shaping filter following the modulator. In this filter the lower sideband is suppressed completely except for those frequencies which are within 500 kc of television carrier. Lower sideband frequencies within 500 kc of carrier are suppressed only partly as also are upper sideband frequencies within 500 kc of carrier to achieve a symmetrical response function in the vestigial sideband region.

It is convenient in a discussion of vestigial sideband transmission to consider the transmission as made up of two components, each symmetrical about carrier frequency, a real or in-phase component and a quadrature component which is a distortion term.[6] The process is illustrated in Fig. 5. Here the response function shown in Fig. 5(a) represents idealized conditions for vestigial sideband transmission. The main sideband is shown extending from carrier frequency Fc to the upper cut-off Fu. A vestige of the lower sideband extends from carrier frequency to the lower cut-off Fv. Constant envelope delay is required in the entire band from Fv to Fu. In the frequency region Fc ± Fv the response characteristic is so shaped that the sum of responses at corresponding frequencies above and below carrier add to a constant value. The summing of signal components in the vestigal bands above and below carrier is accomplished by the receiver demodulator.

The response function of Fig. 5(a) may be considered to be the sum of the two response functions 5(b) and 5(c) which have even and odd symmetry respectively about the carrier Fc. Both 5(b) and 5(c) are double sideband functions. The component in Fig. 5(b) represents the

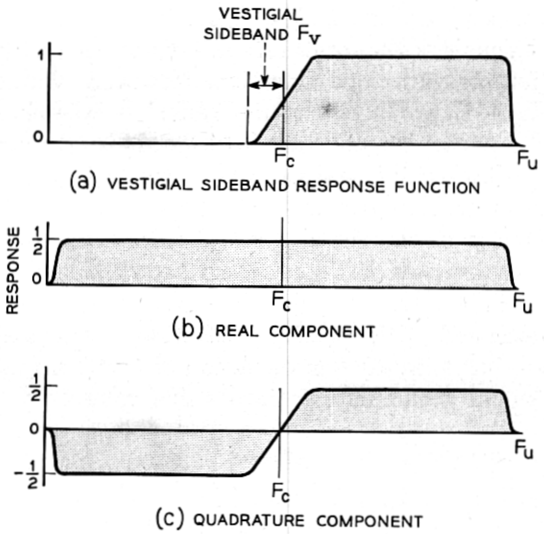


Fig. 5 — Response function of a vestigial sideband carrier system.

normal double sideband output of the modulator supplied with video and carrier signals. The other component, Fig. 5(c), represents the output of a modulator supplied with carrier and signal voltages each shifted in phase by 90° and with the amplitude attenuated as shown in the vestigal region Fc ± Fv. The modulation transmitted by a circuit with the response function of Fig. 5(c) is called the quadrature component and is related to the normal modulation, depending upon the shape and extent of the vestigial sideband. Fig. 6 illustrates the real and quadrature components of an idealized rectangular wave form demodulated after transmission over circuits having the response functions of Fig. 5, respectively.

The 90° shift of carrier frequency in the quadrature component of the vestigial sideband signal makes possible the suppresssion of this component. The transmitted vestigial sideband signal may be written

$$V(t) = P(t) \text{ Cos } ct + Q(t) \text{ Sin } ct, \tag{1}$$

where $c = 2\pi$ times carrier frequency and $P(t)$ and $Q(t)$ are "real" and quadrature modulating functions[6] typically as represented on Fig. 6.

The demodulator may be regarded as an ideal multiplier of signal and carrier supply. Let the carrier supply be denoted:

$$C(t) = \text{Cos } (ct - \phi), \tag{2}$$

where ϕ is the phase of the receiver carrier supply relative to the carrier factor of the "real" component of the signal. The demodulator output is the product $V(t) \times C(t)$. A low-pass filter rejects the output components in the band of frequencies about twice carrier frequency so that the demodulated video signal output is the lower frequency component. Thus, neglecting a factor of ½,

$$V_0(t) = P(t) \text{ Cos } \phi + Q(t) \text{ Sin } \phi. \tag{3}$$

It is seen that real and quadrature video components in the demodulator

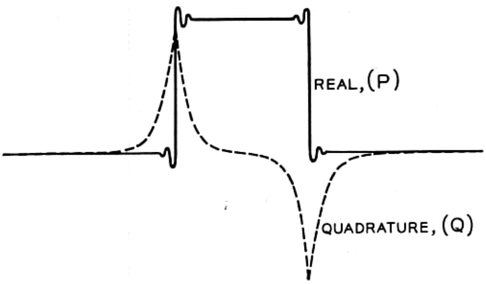


Fig. 6 — Real and quadrature components of a rectangular pulse after vestigial sideband transmission.

output exist in the same proportion as the components of the demodulator carrier supply in phase and in quadrature respectively with the real carrier component of the vestigial signal. By providing carrier exactly in phase with the real component of the signal the quadrature component in the output may be suppressed completely. It has been determined that to suppress the quadrature component resulting from the L3 vestigial band shape to barely perceptible (threshold) values the phase angle of the carrier regenerated at the receiver must be maintained to an accuracy of plus or minus 2.5 degrees. A requirement for one demodulator, when six pairs of terminals contribute to produce quadrature distortion at threshold value, becomes 2.5 degrees divided by the square root of six, or about one degree.

The regeneration of carrier at the receiver is one of the principal L3 terminal features. Here a 4.139-mc carrier must be provided to demodulate the "over-modulated" L3 signal. The required carrier must be reconstituted from information carried in the signal itself. It would be possible, of course, to transmit separately a signal from which carrier frequency could be derived but carrier frequency is really the smallest part of the required information. It is the phase angle of the carrier of the received signal which must be duplicated closely at the demodulator and separate transmission of carrier phase angle does not seem feasible. A phase controlled oscillator is employed for the carrier supply at the receiver, with phase control obtained from information residing in the signal itself and frequency synchronization an additional burden upon the phase control system.

The basis for synchronizing the receiver oscillator to the carrier of the received signal lies in the phase angle of the carrier frequency component of the vestigial sideband signal averaged over a period of time of the order of one frame scanning period. Referring to Fig. 5(b) and 5(c) again, it may be noticed that the quadrature response function is zero at carrier frequency. This means that the quadrature component of the transmitted signal contains no carrier frequency component and will not affect the determination of the real carrier component phase angle based upon averaging over a sufficient period of time. Another signal characteristic presents more serious problems. The degree of modulation employed in L3, shown in Fig. 4(d), makes the average carrier polarity indeterminate. That is, the carrier polarity for a video amplitude corresponding to picture white is opposite to that corresponding to picture black or sync pulses. The polarity reverses as the composite signal changes through its half peak-to-peak value. The average polarity determined from a predominantly white picture is thus opposite to that determined from a predominantly black picture. A carrier oscillator, phase synchronized to the average carrier phase of the signal would execute 180° phase reversals as picture content changed from average white to average black, producing sudden video polarity reversals at the demodulator output.

This signal carrier polarity ambiguity which is momentary in character can be exchanged for one which is not time variable by a multiplication operation. The modulated signal is squared, i.e., multiplied by itself on an instantaneous basis, in a square law circuit. Such an operation squares carrier amplitude and doubles carrier frequency and phase angle, the latter effect converting 180 degree phase reversals into 360 degree changes which are indeterminable in the average phase detector. Under these conditions the phase synchronized demodulator carrier supply, stably locked to the average phase of the squared signal, experiences no phase reversals with change in picture content. The ambiguity now is in the determination of incoming signal polarity. The squaring operation eliminates any basis for determining polarity so that the demodulator carrier may with equal likelihood lock to either polarity relative to the signal and thereby at the demodulator output produce video signal waveforms of either polarity.

The method used to secure phase synchronization of the local receiver oscillator to the received signal is described next with reference to Fig. 7. Signal from the line together with the output from the carrier oscillator are brought to the demodulator where the desired video output signal is obtained as the lower sideband of the modulation product. This process has already been described, equations 1 to 3. In the carrier regeneration process signal and carrier phase shifted by 45 degrees (equations 4 & 5) are each squared in square law circuits.

$$\mathrm{V(t)} = P \operatorname{Cos} ct + Q \operatorname{Sin} \mathrm{ct}, \tag{4}$$

$$C(t) \angle 45° = \pm \operatorname{Cos} (ct - \phi - \pi/4). \tag{5}$$

Band pass filters select from the squaring circuit output signal frequencies in the neighborhood of twice carrier frequency. From the signal squarer,

$$\tfrac{1}{2}(P^2 - Q^2) \operatorname{Cos} 2ct + P\, Q \operatorname{Sin} 2ct \tag{6}$$

and from the carrier squarer,

$$\tfrac{1}{2} \operatorname{Sin} (2ct - 2\phi). \tag{7}$$

The two squared signals at twice carrier frequency are multiplied together in a product modulator. This product contains signals in two bands of frequencies, one band in the region of four times carrier frequency and the other in the video frequency band starting at zero frequency. The lower frequency component of this product is selected by a low pass filter following the product modulator yielding,

$$\frac{(P^2 - Q^2)}{2} \text{ Sin } 2\phi + PQ \text{ Cos } 2\phi. \tag{8}$$

The dc component of the low pass filter output is a suitable control voltage for synchronization and is obtained in the limit as the cut-off frequency of the low pass filter is lowered. Average values as produced by the low pass filters are applied as a frequency control voltage to the carrier oscillator.

The first term in equation (8)

$$\frac{(P^2 - Q^2)}{2} \text{ Sin } 2\phi$$

when averaged is, for small errors in carrier phase angle, proportional to the error angle ϕ. The factor of proportionality is recognized as the difference in mean squared values of the "real" and quadrature modulating functions, P and Q, illustrated typically in Fig. 5. This difference is always positive when the modulating signal contains energy components

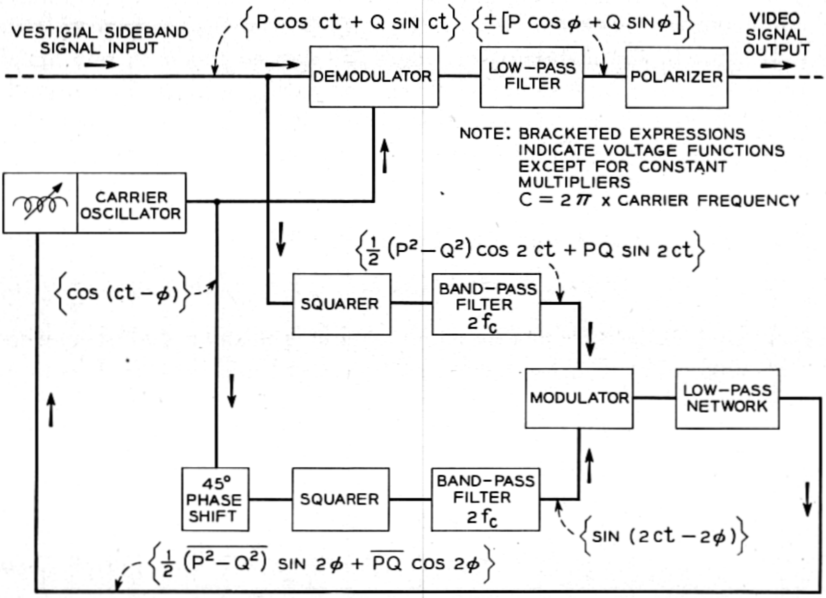


Fig. 7 — Functional diagram of the L3 homodyne demodulation process.

within the bounds of the vestigial sideband since the quadrature response function, Fig. 5(c), is attenuated relative to the "real" response function in this band. In the present case, with a 500 kilocycle vestigial bandwidth and a composite video waveform for a modulating function, the amplitude of Q^2 for control purposes is negligible compared with P^2.

The second term of equation (8), $PQ \operatorname{Cos} 2\phi$, contains no dc component since Q itself contains no dc component and all other frequency components of Q are shifted 90° in phase relative to corresponding components in P. The dc control voltage therefore is not modified by the existence of the second term of equation (8). However, the function PQ does contain sum and difference frequencies due to the cross products of the spectra of P and Q. These frequencies in the control voltage tend to be large compared with corresponding frequencies due to the products P^2 and Q^2 since the trigonometric multiplier Cos 2ϕ, equation (8), is large when the phase angle error is small. The effect of the term PQ Cos 2ϕ is that of phase modulation of the receiver carrier supply and its suppression determines the characteristics required of the low pass filter which averages the control signal. At the penalty of sluggish synchronization and restricted oscillator pull-in range the phase modulation can be reduced to arbitrarily small values. For our purposes a pull-in range of ± 20 cps can be achieved with phase modulation less than ± 0.1 degree with adequate margins.

The control voltage is applied to a tuning element in the receiver oscillator, in this case a small saturable reactor made with ferrite as a core material. This reactor is part of the series resonant quartz crystal circuit which determines the oscillator frequency and is capable of shifting the frequency in response to the control voltage by $\pm$ 20 cps, a figure chosen as safely less than the first sideband components of the transmitted signal which are $\pm$ 30 cycles from carrier frequency. This precaution avoids possible synchronization of the local oscillator to a signal sideband frequency rather than to the carrier.

Sufficient gain is provided in the carrier frequency control loop just described so that the maximum frequency difference encountered between transmitting and receiving oscillators is corrected by the phase control voltage due to a steady state phase angle error, ϕ, less than ½ degree. The control characteristic of the saturable reactor may be expressed,

$$\Delta f = A(P^2 - Q^2) \operatorname{Sin} 2\phi, \tag{9}$$

where Δf is the frequency shift introduced by the reactor and A is the factor proportional to required loop gain.

One other factor to be considered is the stability criterion of the frequency control circuit as a feedback loop.[7] In this case two factors contribute to loop phase shift. First, the phase angle variation of the oscillator output in response to the control voltage is an integration process. The control voltage changes the oscillator frequency and the resulting phase change can be expressed as the integral with respect to time of the frequency shift,

$$\text{phase, } \theta, = 2\pi \int \Delta f dt. \qquad (10)$$

The integration with respect to time introduces a 90 degree "low-pass" phase shift into the control loop at all frequencies. Second, the averaging low-pass filter introduces phase shift in the same direction so that care must be exercised to avoid instability. In this case a phase stability margin of 45 degrees is provided over a wide range of frequency by designing the low-pass filter as a series of resistance capacitance steps of loss. These are staggered in frequency to produce a cut-off rate of 3 db per octave with a phase shift of 45 degrees over a wide frequency band.

The polarity ambiguity resulting from the squaring process has been demonstrated in the derivation of the phase control voltage. In equation (5) the plus or minus designation indicates that either polarity of carrier signal might be assumed without affecting subsequent expressions. However, in the derivation of the output voltage from the main signal demodulation, equation (3), the output signal polarity reverses if the carrier, equation (2), is assumed with reversed polarity. The carrier polarity established at any given time depends largely upon initial phase conditions when signal is applied.

Correct video polarity at the receiver output is established by a new device called a polarizer which follows the demodulator. This circuit recognized video polarity on the basis of standard features in the composite television waveform. The particular features used in this case are the vertical blanking discontinuities expected once each sixtieth of a second and the duty factor of sync pulses. These two characteristics taken together form a sufficient condition for the determination of polarity of any composite video waveform independent of picture content. The polarity, once recognized to be inverted, is corrected.

Harmonic Distortion

A significant consideration in transmission problems is the generation of distortion by the non-linear amplitude characteristic of the transmission apparatus. In the case of video transmission, non-linearity results in the distortion of brightness values of the transmitted picture and presumably will distort chromaticity values of color television signals. With carrier transmission apparatus, in addition to these effects non-linearity produces extraneous interference patterns at harmonics of the carrier frequency. In L3 with a carrier modulated spectrum concentrated near 4.139 mc., the second harmonic distortion of line repeating amplifiers produces a new spectrum concentrated near 8.278 mc, which is demodulated by the receiving terminal to the region near 4.139 mc. This form of distortion is considerably more disturbing in the final picture than a comparable distortion of brightness values and constitutes a limit to transmission signal-to-noise performance.

Advantage is taken of the spectral distribution of energy of television signals to ameliorate somewhat the effects of second harmonic distortion. A pre-emphasis network is employed in the transmitting terminal to accentuate the amplitude of high frequency components of the signal before transmission. At the receiver a restorer network introduces a complementary frequency characteristic to make the overall transmission characteristic constant with frequency, (see Fig. 8). The restorer network, de-emphasizing the high frequency components, likewise suppresses the second harmonic distortion signals. A limit to the amount of predistortion permitted is set by the maximum amplitudes expected of the high frequency picture components particularly in anticipation of a high frequency color sub-carrier in color television systems. Tentatively, the characteristic of Fig. 8 is chosen as a compromise of these factors.

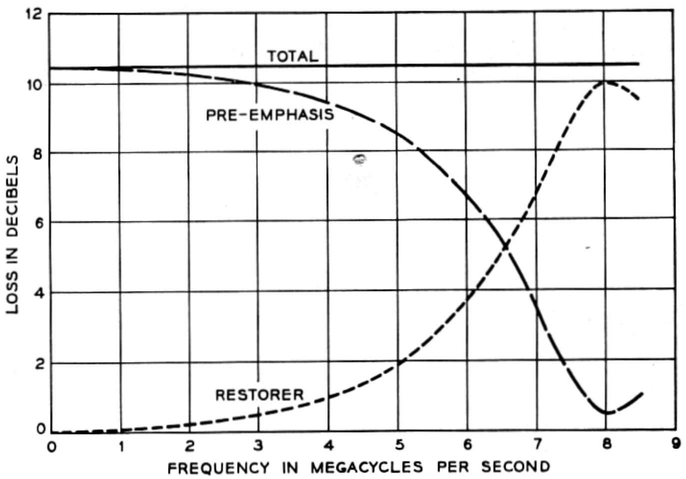


Fig. 8 — High frequency pre-emphasis characteristic.

Filters and Equalizers

A low-pass filter is required before the video signal is modulated, to limit the bandwith of the signal, eliminating possible sources of disturbing cross products, both in the modulator and in the repeaters of the L3 system. This filter has a cut-off at 4.3 mc and provides over 40 db discrimination to all frequencies greater than 4.8 mc. It consists of three m-derived sections and introduces about 1.3 microseconds of envelope delay distortion near its cut-off frequency. This is equalized after modulation by the delay equalizer to be described later.

Following the modulator is the vestigial sideband filter which passes the upper sideband, and provides 60 db discrimination against all frequencies in the lower sideband less than 3.7 mc. The large discrimination is required to avoid interference with the telephone channels during transmission over the coaxial line. This filter provides a controlled loss characteristic to frequencies in the band 3.64 to 4.64 mc which satisfies the requirement for vestigial sideband transmission. The response function for this band is shown on Fig. 2. A flat transmission characteristic including the effective pass band loss of the video LP filter is maintained over the entire upper sideband from 4.6 to 8.44 mc. A second low pass filter after the modulator provides at least 50 db discrimination against third and higher harmonics of the TV carrier (4.139 mc).

A four section high-pass filter designed by the insertion loss method[8] is used to supply 50 db of the discrimination at frequencies less than

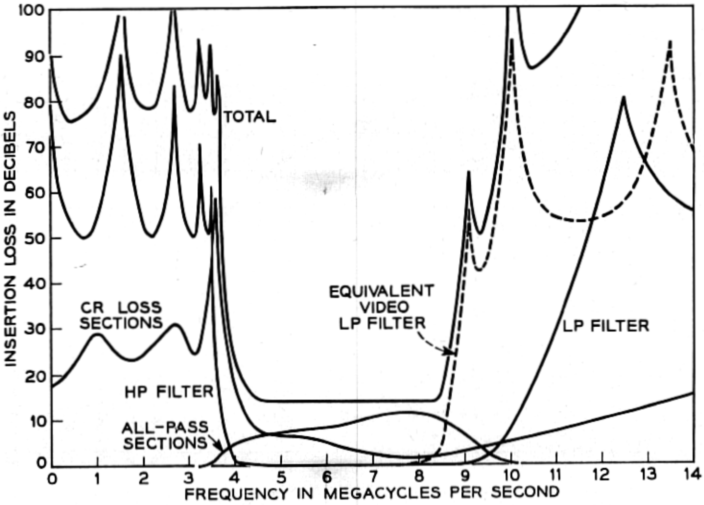


Fig. 9 — Insertion loss of the transmitting terminal filters.

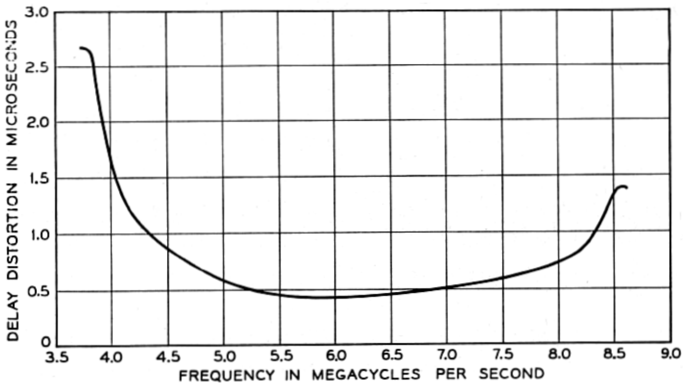


Fig. 10 — Envelope delay distortion of the transmitting terminal filters.

3.5 mc. This filter has a low-pass shunt network at the input to maintain its stopband impedance at 75 ohms. This is required because to produce a uniform frequency characteristic the impedance facing the modulator has to provide a reflection coefficient not exceeding 3 per cent. The remaining discrimination at frequencies less than 3.7 mc and the major part of the vestigial sideband shaping are provided by a group of six constant resistance equalizer sections, as shown on Fig. 9. The low-pass filter to suppress carrier harmonics consists of 2 m-derived filter sections.

The delay distortion of the complete set of filters and loss equalizer sections is shown on Fig. 10. This includes the equivalent delay distortion of the video low-pass filter, translated in frequency for equalization after the modulator. The distortion must be equalized to a constant delay over the television band. A delay equalizer to do this is incorporated in the filter. It was designed by a potential analogue method[9] and consists of 24 all-pass sections, each having the schematic as shown in Fig. 11. A redundant capacitor is used to avoid excessively small capacity values.

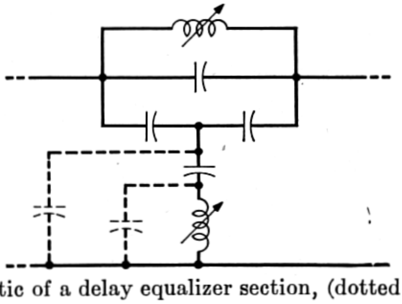


Fig. 11 — Schematic of a delay equalizer section, (dotted capacitors are parasitic).

The capacitors shown dotted are parasitic elements which must be compensated for by modifying the design values of the other elements. This group of sections has an insertion loss varying from 2 to 12 db across the pass band of the filter, caused by the dissipation of the elements. This loss, of course, must be taken into account in the design of the loss equalizer sections.

The objective is to obtain equivalent video transmission through the terminal flat to limits varying from ± 0.02 db to ± 0.10 db, depending upon the frequency characteristic of the deviation. To achieve this, close control of the dissipation loss is essential. The inductor losses which cause the major part of the delay equalizer loss characteristic vary up to ± 15 per cent from nominal values. As a means for controlling inductor Q to ± 0.5 per cent or better, a "Q adjusting screw" is employed. The inductors are solenoids wound on molded tubes with a threaded hole through the center. A threaded magnetic dust core is used for inductance adjustment. Its travel can be limited to the distance from the center to one end of the form without losing adjustment range. By introducing an additional core made of solid magnetic iron into the field of the solenoid, using the opposite end of the form, an adjustment is provided which reduces the Q in a continuous manner as the second screw is advanced into the form. The reduction in Q is caused by the losses in the iron, and normally these would cause a reduction in inductance also. However, the permeability of iron causes an increased concentration of field which tends to increase the inductance. A balance between these tendencies to decrease and to increase the inductance is obtained by controlling the geometry of the Q adjusting core. As a result, a reduction of up to 50 per cent in Q can be obtained, accompanied by a change of less than one per cent in the inductance. Models of the inductor and the adjusting screws are shown on Fig. 12. This adjusting screw in conjunction with the magnetic dust core provides an accurate and economical means for adjusting simultaneously both inductance and dissipation in each inductor of the delay equalizer.

The flat transmission level for the upper sideband and the shaped cut-off for the vestigial sideband were obtained by including loss equalizer sections, assuming Q factors for the all-pass sections of about 20 per cent less than the nominal Q of the inductors. As a final step in the design, the Q factors were modified to absorb in the loss of the all-pass sections the residual loss distortion uncompensated by the loss equalizer sections. This in effect provided the use of 24 additional parameters for shaping the loss in the pass band and resulted in an improved loss characteristic.

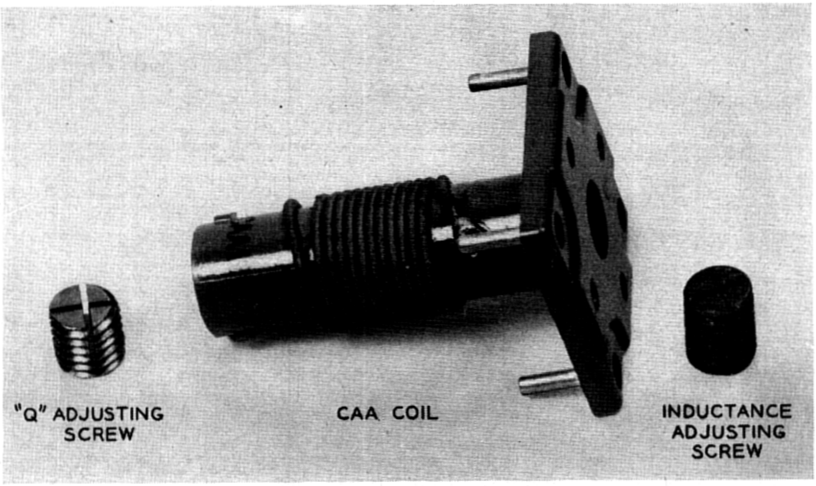


Fig. 12 — Inductor with adjusting screws.

The close limits on delay distortion can be met only by close control of the adjustments on the individual all-pass sections. In order to obtain reproducible results to the order of $\pm 0.1°$ for the phase shift and ± 0.01 db for the insertion loss of the individual sections, a special fixture is employed to make the connection between the section and the measuring circuit. This is shown in Fig. 13. The fixture can be clamped on the network terminals quickly without soldering and provides coaxial patch cords with plugs for connection to the measuring circuit. Each section is mounted in an individual container with shielding between the inductors to reduce coupling. Each inductor is resonated with its associated capacitors and the dissipation is adjusted by adjusting the two cores. The construction of a typical section is illustrated by Fig. 14.

An important consideration in obtaining smooth loss and delay characteristics for the delay equalizer is the reflection coefficient of each section. Poor reflection coefficients cause reflections and interactions between all-pass sections. Due to the large phase slope of the equalizer these tend to produce frequency characteristics with large numbers of loss and delay ripples across the frequency band for which transmission requirements are most severe. Reflection coefficients of 2 per cent or less at all frequencies in the TV band have been obtained for all delay sections by taking the following precautions:

1. Mutual coupling is limited between the two inductors in each section by use of a shield in the section container. As little as 0.1 per cent

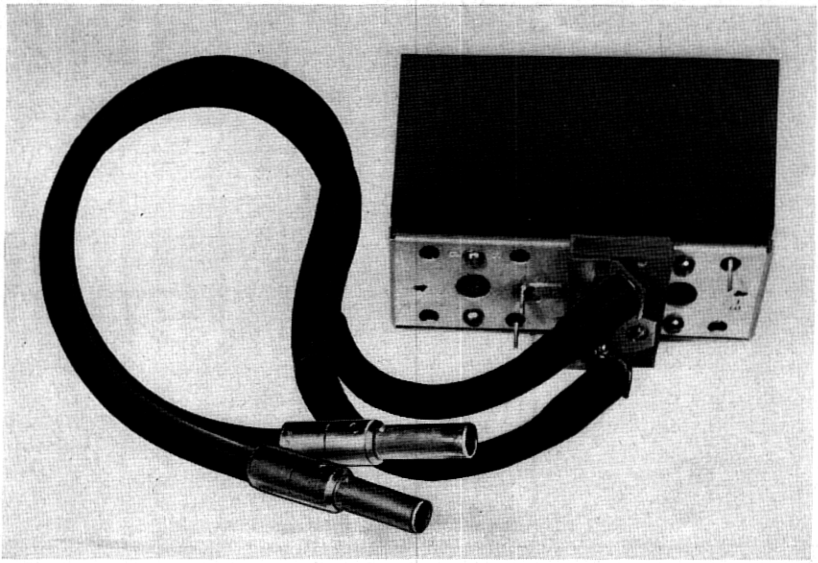


Fig. 13 — Fixture for delay equalizer section adjustment.

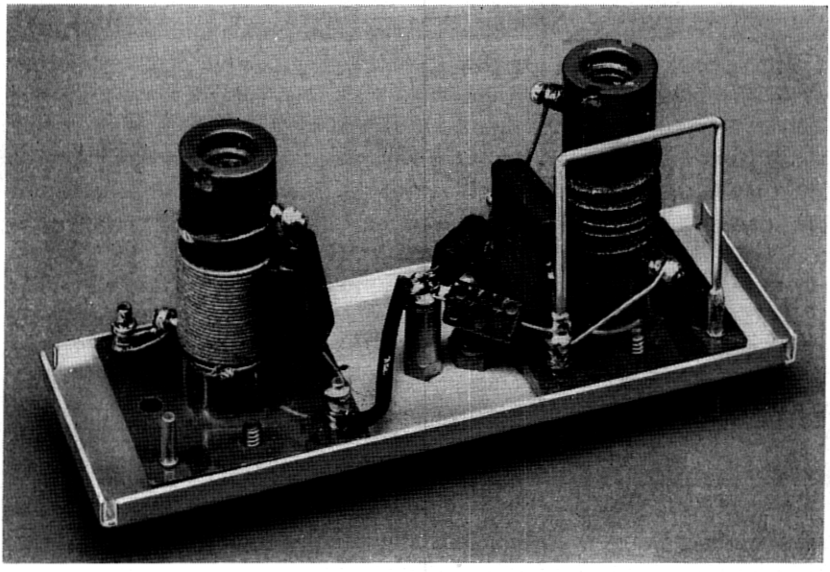


Fig. 14 — Model of a typical delay equalizer section.

coupling coefficient can cause a reflection coefficient of 1 per cent in certain sections at the frequency of 180° phase shift.

2. The Q factors of both inductors are adjusted to be equal in each section.

3. The values of the capacitors are modified to compensate for the presence of the parasitic capacitances associated with the shunt arm, Fig. 11.

The measured insertion loss characteristic and phase shift deviation from linear phase slope for one model of the transmitting filter and delay equalizer is shown on Fig. 15. This has been reduced to video frequency to show the detailed residual distortion which results from the addition of the vestigial lower and upper sideband. The delay equalization is maintained for about 200 kc above the loss cut-off to provide for at least 30-db insertion loss at frequencies where the delay distortion becomes large. Without this precaution the transient response is characterized by a severe "cut-off ring" distortion which is a slowly damped oscillation at cut-off frequency generated by high frequency signal components.

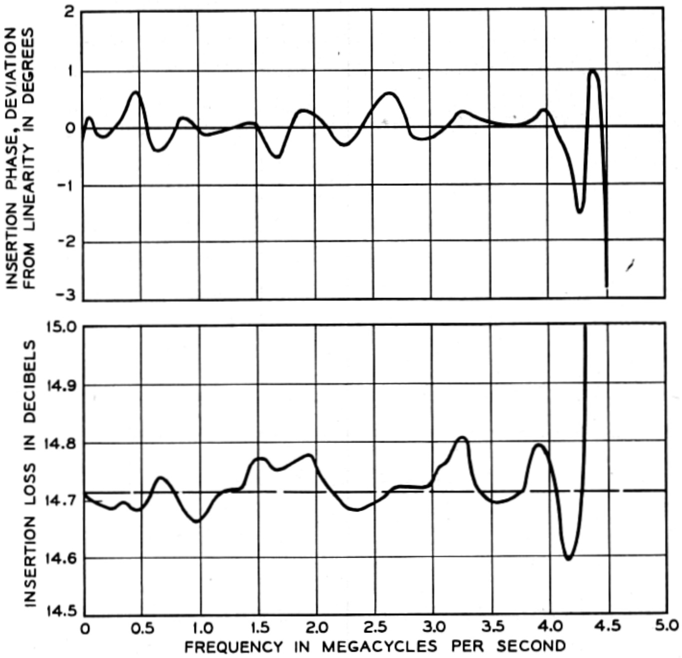


Fig. 15 — Measured pass band performance of a trial model of the transmitting filters and equalizer.

Variations in transmission response in the region near television carrier cause noticeable "smear" distortion on the received images. These variations can be introduced by small changes in element values of the filter or equalizer sections following initial adjustment, or by other changes in both transmitter and receiver. As a means of compensating for such variations, two adjustable loss equalizers have been provided, each with adjustable maximum loss at carrier frequency. One has a half-loss point at 300 kc and the other at 80 kc from the carrier. These are adjusted for minimum "smear" at periodic intervals.

A fixed equalizer is provided in each terminal to compensate for loss and delay distortion other than that in the filters described above. For convenience, this equalization is done at modulated frequencies. It is expected that additional means for periodic re-equalization of the terminal gain and phase characteristics will be necessary to achieve 4,000 mile, tandem terminal transmission quality objectives.

Pilot Frequencies Filters

As described elsewhere,[1] the L3 system has six pilot frequencies for regulating automatically the transmission characteristic of the line. Two of these, 7.266 mc and 8.320 mc, are in the television band. At the transmitting terminal a pilot elimination filter for these two frequencies is required to prevent energy in the TV signal near these frequencies from disturbing the pilot levels on the system. At the receiving terminals, these pilot frequencies must be removed from the TV band to avoid interfering effects in the output TV signal. A discrimination of 50 db for frequencies within 20 cycles of the two pilot frequencies is required. Also at the receiver a carrier elimination filter is required to suppress the residual carrier leak from the demodulator. This requires 30 db suppression to frequencies within 20 cycles of 4.139 mc.

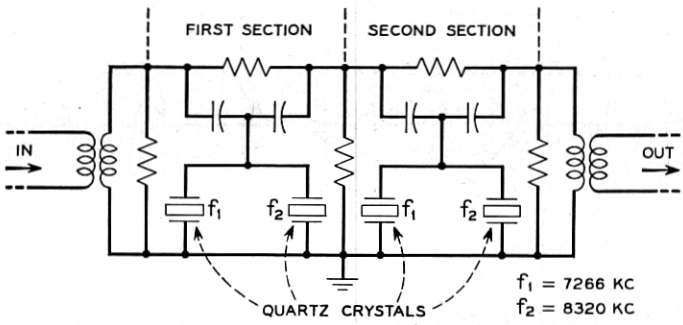


Fig. 16 — Simplified schematic of the pilot frequency band elimination filter.

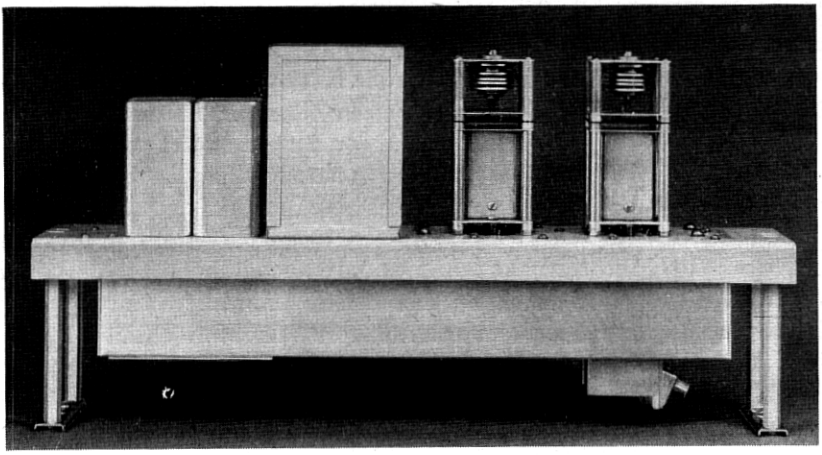


Fig. 17 — Model of the carrier suppression filter.

To avoid removing excessively broad bands of frequencies from the television band, and thereby generating visible distortion in the transmitted picture, narrow bandwidth crystal filters are used for both of these applications. The pilot elimination filter has its 3 db loss points at about 1,000 cycles on either side of the pilots. The carrier suppression filter has its 3 db loss points at 150 cycles on either side of the carrier. In the past, spurious or secondary responses in the crystal units could not be limited to sufficiently low values to permit available crystals to be used in broad band circuits without elaborate means for suppressing the unwanted responses. In the L1 system, for example, a balancing circuit using hybrid coils was employed for this function. Techniques for reducing unwanted responses in the crystal units by special contouring of the blanks and by precise optimum area plating have been developed recently.[10] The use of these methods has resulted in crystal units in which the unwanted responses are reduced to the extent that relatively simple filters can be used for these applications. A simplified schematic of the pilot elimination filter is shown on Fig. 16. Small ovens are provided to maintain close temperature control for the quartz cyrstal elements in order to stabilize the resonant frequencies. Mechanical arrangements are illustrated on Fig. 17.

Power Equipment

Primary power for the television terminals is 60-cycle ac. It is derived from the L3 motor-alternator power equipment at main repeater stations or, alternatively, from commercial 60-cycle sources. When commercial power is used directly a provision is made for switch-over to emergency power in the event of power failure. The emergency power supply comprises a 130-volt battery, an inverter, and automatic switch-over equipment. If during service the commercial power source fails a switch-over to emergency power is made automatically in a time under one-tenth second.

Dc and heater power provisions involve two departures from previous practice. First, the rectifiers which supply anode and other dc power requirements are non-regulated metallic, (selenium), rectifiers. These are supplied from magnetic ac line voltage regulators to obtain suppression of line voltage variations. The combination of the magnetic ac regulator and the metallic rectifier provides adequate suppression of hum and line voltage variations with, it is expected, greater reliability than alternative vacuum tube regulated rectifiers. Secondly, the magnetic ac line voltage regulator together with an improved heater transformer design make possible the operation of the vacuum tubes at reduced heater voltage to obtain increased thermionic life expectancy. This advantage can be taken only if close control of heater transformer output voltage is possible. In this case, variations in output voltage expected to occur due to temperature variations of the transformer windings substantially have been eliminated by incorporating thermistor temperature compensation in the transformer primary circuit.

T3 Transmitting Terminal

A block diagram of the components of the transmitting terminal is shown in Fig. 18. The composite video signal is received from the Bell System television operation center over 124-ohm balanced cable and

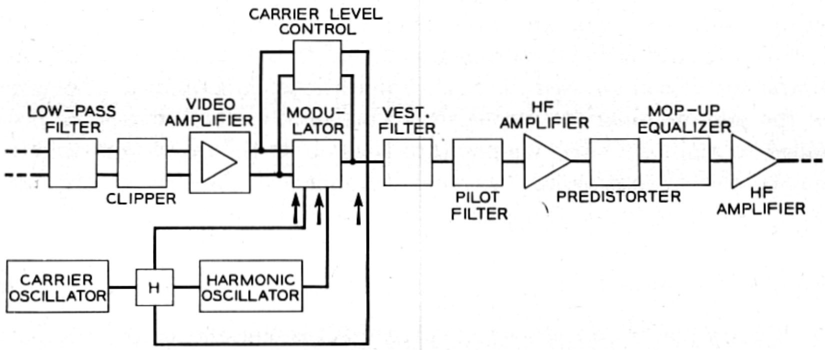


Fig. 18 — L3 transmitting terminal block diagram.

the modulated output signal is delivered to the L3 line facilities over 75-ohm coaxial cable.

An input low-pass filter and a video clipper circuit place ceilings on the maximum transmitted signal bandwidth and signal amplitude, respectively. The clipper circuit is required to protect the 600 telephone channels from inadvertant overloads due to excessive television signals. Normal amplitude signals are not affected by the clipper.

Following are a video amplifier and the modulator together with the required carrier supplies and the carrier level control. This latter device accurately regulates the peak carrier magnitude in the output signal and thereby preserves the desirable maximum modulation.

Following the modulator is the vestigial band filter which, together with its delay equalizer, shapes the double sideband modulator output into the vestigial transmission band. The flat loss of these networks requires that amplification be provided in the carrier frequency band to increase the signal level. The first of two high frequency flat gain amplifiers in the transmitter restores the signal amplitude.

The pilot band elimination filter is next provided to remove television signal energy and other possible interference from particular frequencies allocated to the L3 line pilot signals.

At this point occur the predistorter network for pre-emphasizing the high frequency signal components and a mop-up equalizer. The mop-up equalizer is to provide means for periodic correcting of the transmission characteristic. The high-frequency amplifiers, particularly, change their transmission characteristic as the vacuum tubes age. An additional high frequency amplifier is provided to recover the loss of the foregoing networks and deliver a proper signal level to the line equipment.

R3 Receiving Terminal

The receiver demodulates the signal as transmitted over the L3 line facilities to recover the video signal for transmission over local video circuits. As has been discussed the demodulation is a homodyne process utilizing a local carrier regenerated from information contained in the transmitted signal.

A block diagram of receiver components is shown in Fig. 19. The first components are a group of networks; the restorer to compensate for transmitter predistortion, the fixed and variable mop-up equalizers to compensate for deviations in the receiver frequency characteristic, and a pilot elimination filter to remove L3 pilot frequencies from the signal. A high frequency amplifier provides amplification to compensate for the network losses.

The amplified signal is split into two branches by a hybrid transformer. Part of the signal is carried to the demodulator for detection to video frequencies while the remainder is employed in the carrier regeneration apparatus. The carrier supply oscillator output likewise is split between the same two circuits; the portion supplied to the demodulator providing the carrier energy for demodulation and the part sent to the carrier regeneration circuits providing means for comparison with the input signal. The carrier regeneration equipment comprises mainly the phase comparison circuit together with means for changing electronically the frequency of the carrier supply oscillator.

The phase synchronized carrier supply together with the associated third harmonic supply provide demodulating carriers which translate the carrier spectrum down in frequency to the original video spectrum. The demodulator output also contains an unwanted upper sideband in the 8 to 12-mc region. This signal and residual carrier leak from the demodulator are removed from the detected signal by a low-pass filter and the quartz crystal narrow band rejection filter tuned at carrier frequency. Finally the signal is amplified at video frequencies to provide moderately high video frequency transmission levels and transmitted through the reversing relays of the polarizer to the receiver output.

Fig. 20 is a photograph of two equipment bays each eleven feet high which comprise at the left the T3 transmitting terminal and at the right the R3 receiving terminal.

Acknowledgement is made for the many contributions made by our associates toward the successful development of these terminals. These include both ideas related to transmission processes and important contributions in the development of many new circuit components which were necessary to the achievement of the quality and reliability objectives.

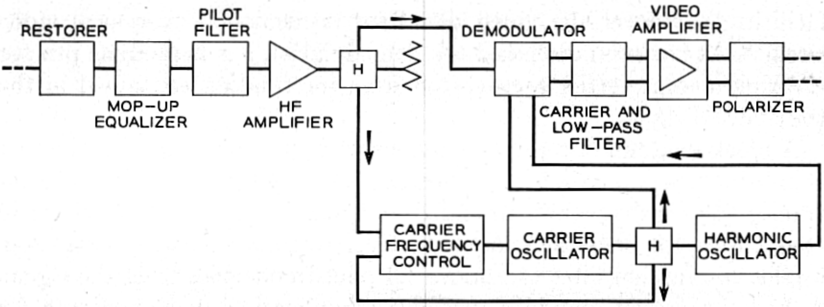


Fig. 19 — L3 receiving terminal block diagram.

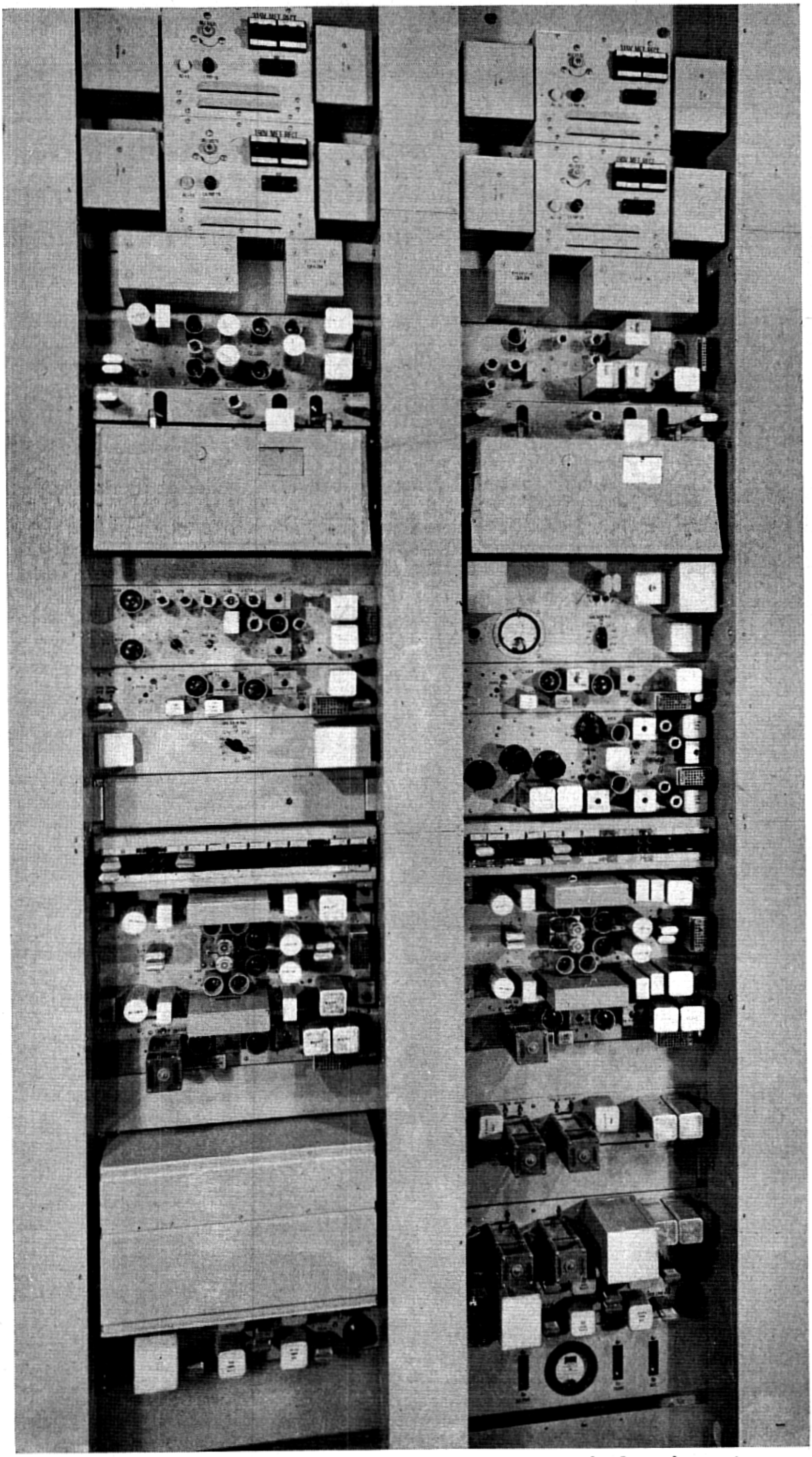


Fig. 20 — Model of transmitting and receiving terminal equipments.

REFERENCES

1. C. H. Elmendorf, R. D. Ehrbar, R. H. Klie, A. J. Grossman, The L3 Coaxial System—System Design, see pp. 781 to 832 of this issue.
2. L. H. Morris, G. H. Lovell, F. R. Dickinson, The L3 Coaxial System-Amplifiers, see pp. 879 to 914 of this issue.
3. L. W. Morrison, Jr., Television Terminals for Coaxial Systems, A.I.E.E., Trans. **68**, pp. 1193–1199, 1949.
4. R. S. Caruthers, Copper Oxide Modulators in Carrier Telephone Systems, Bell Sys. Tech. J., **18**, pp. 315–337, April, 1939.
5. L. A. Meacham, Bridge-Stabilized Oscillator, Bell Sys. Tech. Jl., **17**, pp. 574–591, Oct., 1938.
6. H. Nyquist, Certain Topics in Telegraph Transmission Theory, A.I.E.E., Trans, **47**, pp. 617–644, 1928.
7. H. W. Bode, Network Analysis and Feedback Amplifier Design, D. Van Nostrand Co., 1945.
8. S. Darlington, Synthesis of Reactance Four-Poles, J. of Math. and Phy., **18**, pp. 257–353, Sept. 1939.
9. S. Darlington, The Potential Analogue Method of Network Synthesis, Bell Sys. Tech. J., **30**, pp. 315–365, April, 1951.
10. L. J. LaBrie and D. F. Ciccolella, High Frequency Crystal Units for Use in Selective Networks, submitted for publication, I.R.E. Trans. (Presented at I.R.E. subcommittee meeting, Washington, D. C., Dec. 4, 1952).