The L3 Coaxial System

Amplifiers

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The line amplifiers for the L3 coaxial system are designed to compensate for the loss of the four miles of cable which separate the repeaters; the flat amplifiers are used to compensate for equalizer loss and as transmitting amplifiers. The two types are basically similar, consisting of two feedback amplifiers in tandem separated by an inter-amplifier network; in the line amplifier this network is variable and is automatically adjusted to compensate for variations in cable temperature and for small deviations from the nominal four-mile spacing.

Coupling networks employing high-precision transformers are used to connect the amplifers to the coaxial cable through the required power separation filters. The low impedance windings of the transformers are center-tapped and a balancing network provided in order to match the cable impedance over the transmitted frequency band. The amplifiers are equipped with plug-in tubes of high figure of merit which were developed for this application. A double-triode output stage is used to obtain improved system signal-to-noise performance. Provision is made for preventive maintenance of vacuum tubes and for a controlled adjustment of gain on an in-service basis.

All important components of the amplifier are subject to quality control procedures to assure that the average gain of groups of amplifiers will be held within narrow limits and that individual amplifiers will form a normal distribution around the average. This approach is essential in order to meet system equalization and signal-to-noise objectives. Careful mechanical design and rigid control of the mechanical aspects of manufacture are necessary to minimize gain variations which might be caused by variations of parasitic circuit elements and unwanted feedback effects. Special measures were required to keep the temperature rise within the sealed die-cast housing within tolerable values.

Introduction

In a transmission system such as the L3 coaxial, the degree to which system objectives are achieved is largely dependent on the quality of the amplifiers which compensate for the cable loss. To a considerable extent the same statement applies to the similar flat-gain amplifiers used to make up for the loss of the equalizers at various points along the route. The development of an amplifier which would meet the exacting requirements of the L3 system was in turn dependent on new developments in the fields of vacuum tube design and circuitry, network design techniques, element and network fabrication, and statistical quality control. To these new tools were added the lessons learned in years of manufacture and operation of the preceding L1 system.

The importance of some of these factors can best be illustrated by examining the implications of the amplifier requirements which follow from the material in the companion papers on system design[1] and equalization.[2] Obviously the figure of merit, modulation coefficients and life of the vacuum tubes will be determining factors in setting the amount of feedback that can be obtained and the signal-to-noise ratio of the system. It is not so immediately apparent that system requirements could not be met with the present 4-mile repeater spacing if it were not for the use of quality control at every stage of manufacture from elements, and even the raw materials entering into components, to complete amplifiers. As the companion papers show, however, the equalization plan of the system is predicted on a degree of reproducibility of amplifier gain and delay characteristics obtainable only by quality control applied at every stage of the manufacturing process.

The present equalization plan is based on the assumptions that the gain of the average line amplifier will match the loss of the preceding line section to within 0.15 db, and that the average amplifier gain will not vary from one batch of new amplifiers to another by more than about 0.06 db. Under these assumptions a system equalization plan can be worked out which results in reasonable spacing between equalizers, a tolerable signal-to-noise penalty due to misalignment and equalizer loss, and a practicable procedure for adjusting long systems. Any gross departure from the basic assumptions as to reproducibility of amplifiers would seriously compromise these objectives, which even now are achieved only by using equalization which requires a flat gain amplifier for every four or five line amplifiers. Now it turns out that with the most precise elements that can be made, the gains of individual amplifiers will vary by about ±0.6 db. We need, therefore, a tool which will permit us to control the gain of the average amplifier to an order of magnitude greater precision than we can economically control the individual. This is exactly the effect which modern methods of statistical quality control aim at achieving, and since the entire amplifier is merely the sum of its parts, quality control must start at the roots of the manufacturing process. In order to apply quality control intelligently, and to be sure that all important causes of gain variation are understood, it has been necessary to carry out, side by side with empirical laboratory work, a program of computing the insertion gain of the amplifier, starting with fundamental element values. These computatuions have also proven of value in obtaining satisfactory stability margins in the design of the low and high frequency cut-offs of the feedback loops, and in obtaining preliminary information on amplifier gain deviations for use in equalization planning.

The severe gain and delay reproducibility objectives also have their effect on the mechanical design of the amplifier. At first sight the unit appears to be a lumped constant structure rather than one in which distributed effects would be of paramount importance. Usually the circuit designer in such a case is interested in the mechanical design only for reasons of neatness and economy, but when we look deeper we find that in the amplifier structure as a whole as well as in the case of certain components, the effects of distributed capacity and inductance could easily defeat our objectives if the mechanical design were not such as to assure precise control of element placement and wiring lengths.

For these reasons, an order of mechanical accuracy is specified beyond that which can be justified by the accuracy of transmission measurements on individual amplifiers. This striving for mechanical accuracy is carried all the way through, from element piece parts through subassemblies to the assembly on the final amplifier framework. The logical expectation is that by reproducing the amplifiers as exactly as possible we will control not only the known elements but also the many parasitic effects, and thereby minimize the appearance of shifts in amplifier gain and delay which would be substantial in the system even though difficult to detect in the measurement of individual amplifiers.

Another design feature of the amplifier based on the system equalization point of view is the omission of all adjustable elements, or trimmers, to control the transmission. By eliminating gain adjustments, the possibility of adjusting one element to compensate for the short-comings of another, and the possibility of systematic errors of setting due to faulty or inaccurate adjustment techniques, are both eliminated. Either of these possibilities would tend to convert relatively large random effects into smaller but systematic effects, a conversion which would penalize the system equalization. Obviously, however, the elimination of gain adjusting elements puts a higher premium than ever on the requirement that all elements, including tube capacities, show only small random variations about tightly controlled nominal values.

In addition to the gain requirements discussed above, the following line amplifier objectives are to be met:

1. The gain of the amplifier must be continuously variable, under the control of a regulator circuit, to compensate for differences in lengths of cable sections and for variations of cable loss caused by temperature changes. The shape of the gain change should match the square root of frequency shape of the line loss change over the transmitted band to within a few hundredths of a db. This accuracy of shape, which of course is based on equalization considerations, should hold over the entire regulation range, which is about ± 6 db at the top transmitted frequency.

2. The input and output impedances of the amplifier should match the cable impedance in order to minimize the effects, on television transmission, of echoes caused by line irregularities such as splices. Tolerable values of reflection coefficient at amplifier input and output are about 5 per cent at the television carrier and 10 to 15 per cent at upper band edge.

3. Feedback consistent with system modulation requirements, shaped across the transmitted band to minimize low frequency intermodulation products and to give a smooth, easily equalized shape of gain change as tubes age, must be obtained while maintaining adequate margins against singing.

4. Other requirements include design and selection of elements to assure as small a change of gain versus ambient temperature as possible, mechanical design provisions for keeping the temperature rise within the unit to a minimum both in order to keep the temperature-gain effect small and to obtain long element life, a sealed housing to avoid damage by humidity in exposed locations, and provision of facilities for testing tubes in service.

Configuration

The circuit configuration of the line amplifier is shown in Fig. 1. It consists of two independent feedback amplifiers with a regulating network between them acting like a two-terminal interstage. Each of the amplifiers is essentially a two-stage circuit — the input amplifier literally so and the output amplifier essentially so since the double-triode output circuit acts like a single stage. Each amplifier is connected to the coaxial through a coupling network which consists of a transformer

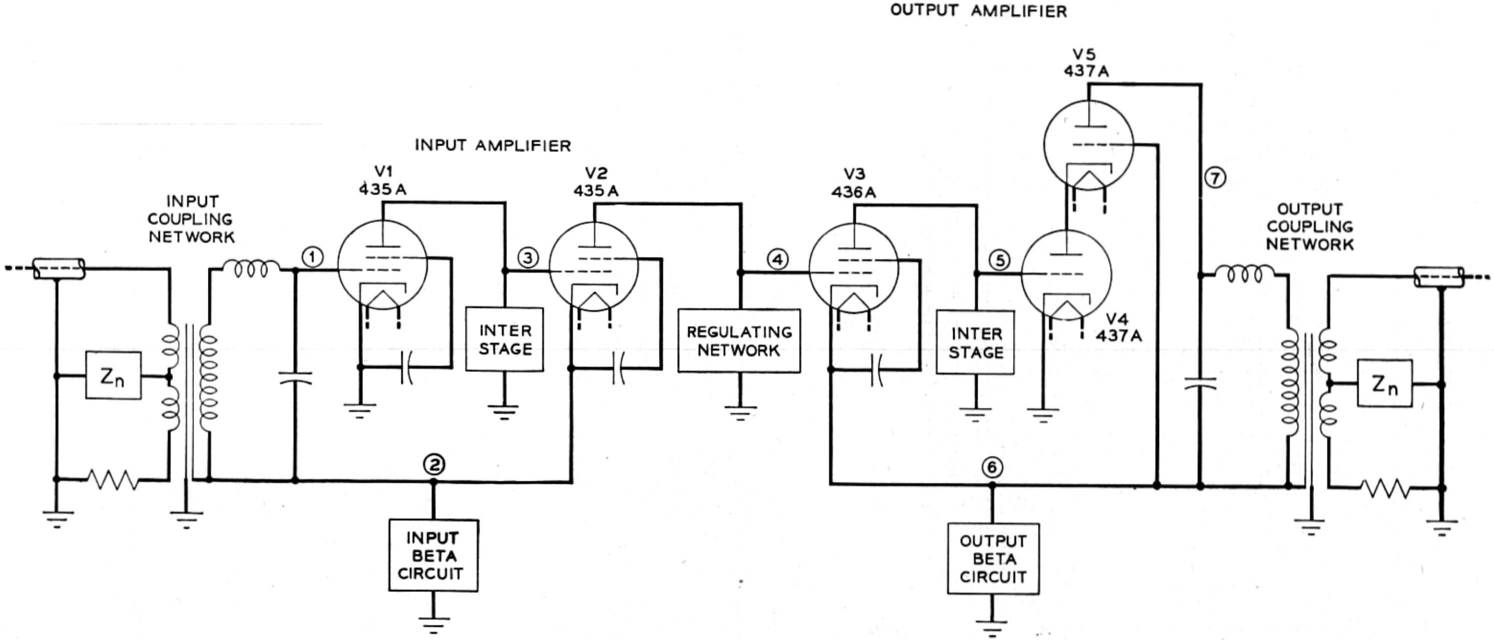


Fig. 1 — L3 line amplifier configuration.

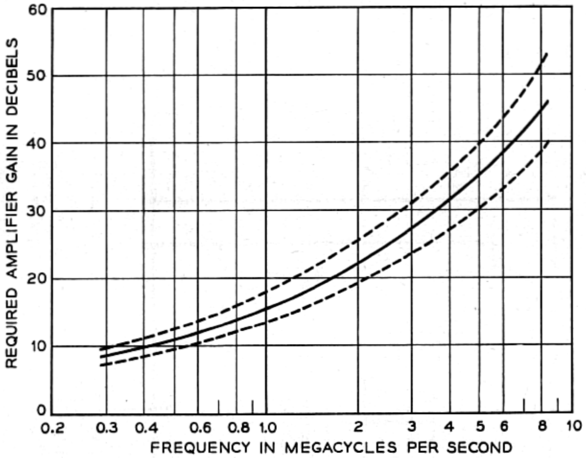


Fig. 2 — Required line amplifier gain.

plus gain shaping and impedance adjusting elements. Since there is no feedback around the coupling networks, they directly affect the insertion gain of the amplifier, as do the two beta circuits and the regulating network. The required shaping of insertion gain across the transmitted band is obtained and controlled by the design of these five networks. Fig. 2 shows the required amplifier gain for nominal and extreme thermistor settings; these required gains differ from the line loss by the small losses of the associated repeater components. At mid-range thermistor setting about 37 db of gain shaping is needed. The manner in which this shaping is distributed among the five networks has important effects on the feedback which can be obtained and the sensitivity of amplifier gain to element variations.

This configuration offers several advantages, one of the most important of which is that the regulating network is between the amplifiers. In most other configurations, the only gain-determining network available for the regulating function is the beta circuit. When the feedback is not infinite, this introduces errors for which it is difficult to compensate, and limits the available feedback by complicating the design. In this configuration, the impedances which the amplifiers effectively present as shunts across the regulating network are high and can be allowed for in the design so that the regulation error can be made much smaller than the error associated with beta circuit regulation in an amplifier having relatively little feedback. Other major advantages are the superior signal-to-noise performance and the relative simplicity of the feedback loops of this amplifier as compared to alternative designs.

Mechanical Design

The mechanical assembly of the amplifier, like the configuration, is divided into three main sections. The input and output amplifiers are mounted on separate chassis which are designed for ready removal from the main base casting. The regulating network is mounted in an enclosed, shielded compartment which also serves to shield the component amplifiers from each other. Figs. 3 and 4 show the assembled amplifier.

Because of the wide frequency range and close control of parasitic capacity and lead inductance required, the L3 amplifier was designed as an integrated whole, and all networks were designed with, and as part of, the amplifier. In each case circuit elements were placed in space in the best possible position for optimum electrical performance, and supporting structures were then designed to maintain the desired space relationships. These supporting structures are made as separate units which mount on the amplifier chassis, so that the networks can be individually tested before final assembly into the amplifiers, and can be removed for repair of replacement if necessary.

Heretofore, this method of design has been impracticable because it results in very complicated supporting structures. It was feasible in this case because of the availability of a new type of material, which removes many of the mechanical design constraints. This is a cold casting resin, and parts are produced in a cheap phenolic mold. Since the process is practically equivalent to the sand-casting of metals, complex parts can be economically manufactured even in the relatively small quantities required for L3 production. Where necessary to assure accurate

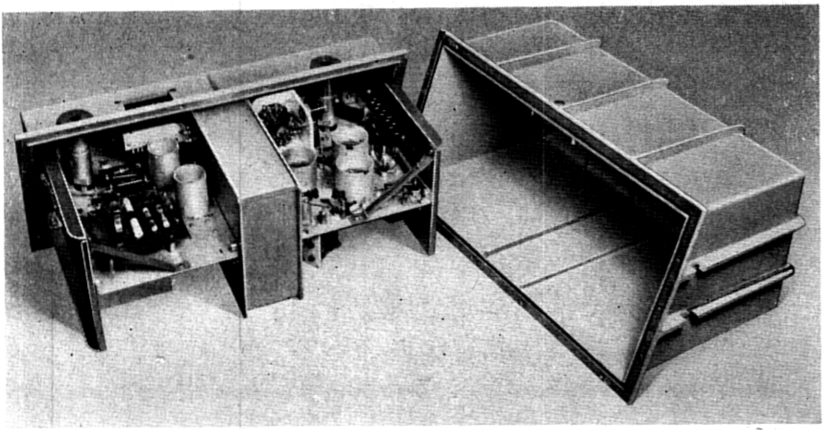


Fig. 3 — Line amplifier and housing.

location or orientation of parts, the cast resin frameworks are milled or spot-faced. The structure used in the input interstage, shown in Fig. 5, is an example. Circuit elements are wired to pins driven into the casting or, in some cases, to wires imbedded in the material. All wiring can thus be made direct with a minimum of bending and no doubling back, resulting in a reproducible and uniform product.

The entire amplifier is housed in a sealed die cast container to protect the components from humidity damage. Dessicant is enclosed in each amplifier. It would have been desirable to make this housing of aluminum, but the high melting point of this metal makes it impractical for such a large casting to meet the air-tightness requirement. A zinc-base die casting alloy was therefore used. Sealing is accomplished by rubber gaskets at all openings for connections and at the joint between the two parts of the housing. The removable part of the casting which serves as a cover is made as large a part of the total housing as possible, in order to provide maximum accessibility for maintenance of the units mounted on the base. The entire housed unit is arranged to mount on a relay rack mounted panel by means of slides which are self-locking. Signal and power connections are made by means of flexible cords which

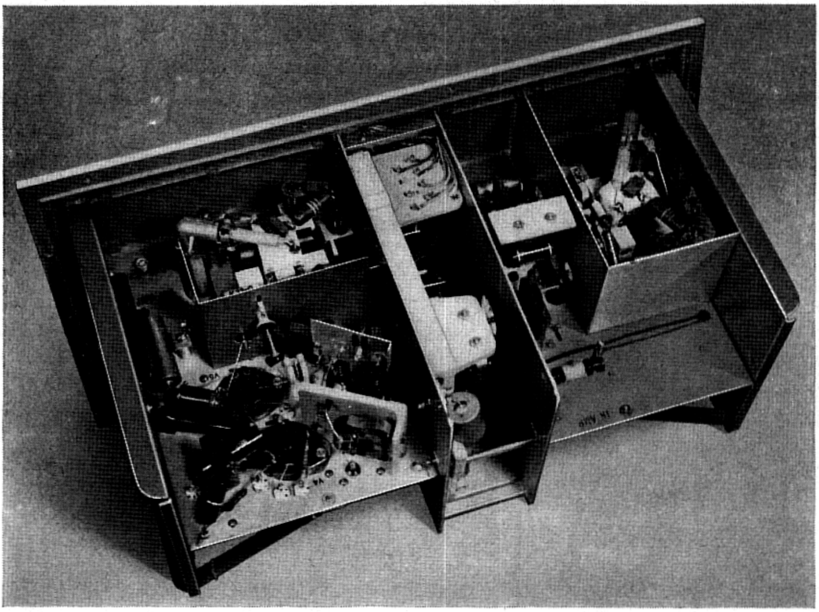


Fig. 4 — Line amplifier, wiring side.

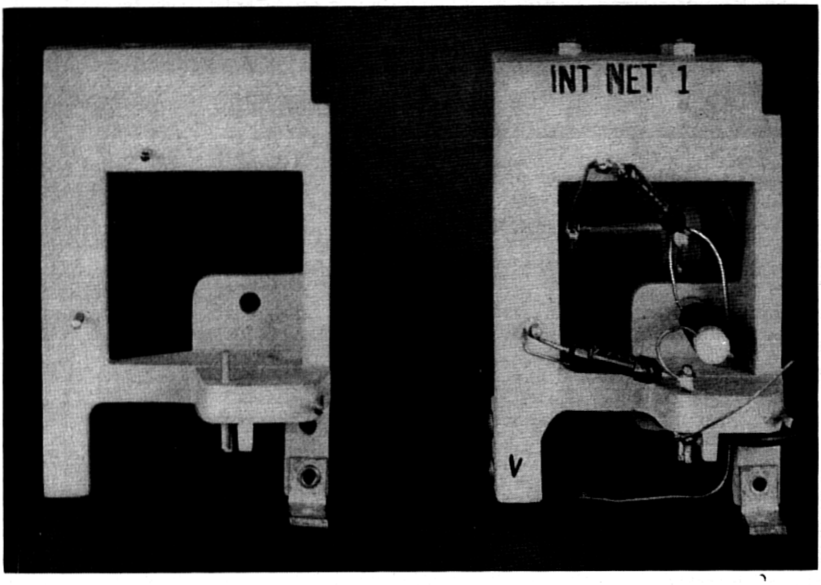


Fig. 5 — Input interstage, illustrating use of case resin frameworks.

are available on the panel to plug into the unit after it is in place on the slides.

Vacuum Tubes

The tubes have been fully described in an earlier paper;[3] their characteristics are summarized in Table I. They are plugged into conventional sockets, and single rather than parallel tubes are used in each stage. These are departures from the practice of the L1 coaxial system, in which two tubes in parallel were soldered in each stage. The use of sockets increases the parasitic capacities and reduces the obtainable feedback by one or two db, but it was felt that the resulting maintenance economy was worth this sacrifice. With single tubes, the failure of one tube results in a line failure and a switch to the protection line. At first sight, it would appear that using parallel tubes in each stage should greatly decrease the probability of line failure. A study of L1 experience, however, showed that most tube failures could either be forestalled by preventive maintenance, or else were of such a nature (for example, shorts within the tube) that the parallel tube would not afford protection against line failure. The reliability advantage of parallel tubes, then, turns out to be small; their use, on the other hand, increases the wiring

TABLE I — TUBE CHARACTERISTICS, DESIGN CENTER VALUES, NEW TUBES

	435A Tetraode	436A Tetrode	437A Triode
Heater voltage	6.3	6.3	6.3 volts
Heater current	0.3	0.45	0.45 amperes
Plate-cathode voltage	180	180	150 volts
Screen-cathode voltage	150	150	— volts
Grid-ground voltage	9.0	9.0	9.0 volts
Cathode bias resistor	630	315	262 ohms
Plate current	13.1	23.4	40.2 milliamp
Screen current	3.2	8.6	— milliamp
Transconductance	16.5	32.0	47.0ma/volt
Plate resistance	—	—	970 ohms
Grid-cathode bias	1.3	1.1	1.5 volts
Capacitances (Approximate hot values in mmf)			
Grid to plate	0.02	0.04	3.6
Grid to cathode and screen	9.2	18.2	16.4
Grid to heater	0.6	0.8	0.1
Plate to cathode and screen	1.0	1.5	0.7
Heater to cathode and screen	5.4	7.2	5.4
Heater to plate	1.7	1.9	0.2
*Modulation**			
Second order (2F)	36	—	36 db
Third order (3F)	66	—	67 db
Third order "effective" (3F)	60	—	60.5 db
Equivalent noise resistance, ohms	460	180	—

* Ratio of product to fundamental at output for a 0.1 volt rms signal from grid to cathode.

complexity greatly when sockets are used. The small gain in reliability would not compensate for the degradation in performance caused by the complication of wiring and the resulting capacity penalties.

The applied voltages and current drains are shown in Table II.

HEAT DISSIPATION

In common with many modern designs with high gain vacuum tubes, the problem of heat dissipation was acute in the L3 amplifier. The amplifier is enclosed in a sealed housing of sufficient size to dissipate readily the heat generated. However, with the usual types of chassis and tube shield construction, vacuum tube envelope temperatures were so high that long tube life could not be assured. Consequently a new type of tube shield was developed. This shield is of heavy copper tubing equipped

Table II — Power Requirements of Amplifier Unit

6.3 volts (grounded)	1.5 amperes
6.3 volts (190v off ground)	0.45 amperes
+100 volts regulated	0.5 milliamps
+190 volts	68 milliamps
+315 volts	41 milliamps

with internal helical springs mounted in such a manner that each turn of the spring makes contact with both the glass tube envelope and with the copper tube as shown in Fig. 6. Thus each turn of the spring provides two metallic heat conducting paths to carry away the heat from the tube envelope. A total of 480 conducting paths are provided for each of the large tubes used in the output amplifier by the use of these springs. Good contact is made between the copper tubes and the chassis, so that heat generated within the vacuum tube is efficiently conducted through metallic paths to the copper chassis of the amplifier. In order not to raise the temperature of the chassis and consequently the temperature of the circuit components, large ribs are provided in the housing, with which the chassis make intimate contact. A continuous metallic con-

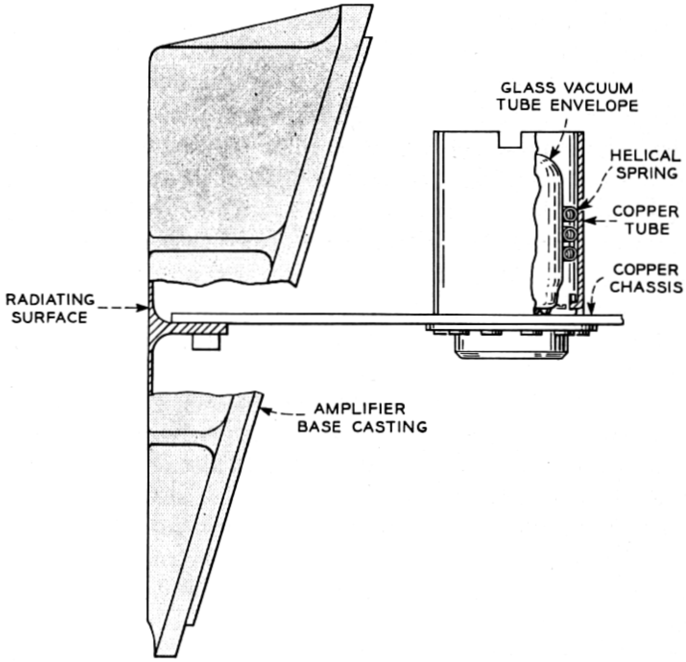


Fig. 6 — Heat conducting tube shield.

ducting path is thus provided from vacuum tube envelope to amplifier housing. This design resulted in a temperature drop in the output amplifier tubes of about 70° C without any temperature rise in the chassis or circuit elements over that produced when ordinary types of shields were used. The smaller vacuum tubes in the input amplifier do not ordinarily run as hot as those in the output amplifier and a temperature drop of only about 55° C was attained by these methods.

Coupling Networks

The input and output coupling networks are essentially identical. The low side of each transformer is a balanced center-tapped winding which together with the balancing network acts as a hybrid, to produce a good 75-ohm impedance facing the cable. The use of this type of connection gives a signal-to-noise advantage over the use of a brute-force high-side terminating network. The advantage is theoretically 3 db in the case of the output coupling network and would approach the same figure in the case of the input network if tube noise were dominant over the resistance noise of the cable.[4] Aside from the fact that the design of a high-side shunt termination network for an off-ground peaked transformer is well-nigh impossible, the use of a balancing network in a hybrid connection has the important additional advantage that the adjustment of this network to obtain a good reflection coefficient has negligible effect on the insertion gain of the circuit.

A relatively modest share of the total shaping required has been allocated to the coupling networks: 5.5 db each, or 11 db total. One reason for this is that although these networks are outside the feedback loops in the usual sense, nevertheless the impedances which they present to the amplifiers are important factors in the feedback design, and the effects which they produce must not be allowed to become so severe as to limit the feedback to too low a value. It is obvious from inspection of Fig. 1 that only a part of the voltage developed across the input beta circuit by the plate current of the second tube will appear as a grid-cathode voltage to drive the first stage. The proportion of the beta circuit voltage which will be thus effective in producing feedback around the loop will be dependent on the potentiometer division between the impedance of the coupling network and the grid-cathode impedance of the first tube. The greater the peaking of the input coupling network, the greater its impedance at high frequencies where the grid-cathode capacitive impedance is already decreasing, and hence the greater the potentiometer term loss. A similar loss occurs in the output amplifier. The plate current of the output stage divides between the output coupling network and the parasitic admittances to ground. The portion of the plate current which returns directly to cathode (ground) through the latter path, without passing through the beta circuit, is not effective in producing loop feedback.

A second and even more important limitation is that the sensitivity of the insertion gain to variations in the coupling network elements is increased as the slope is increased. The same considerations lead us to keep not only the slope but also the gain level or efficiency of these networks relatively low. The maximum possible coupling network gain which can be obtained over the entire frequency spectrum is limited by the capacity across the circuit, as shown by Bode's Resistance Integral Theorem. This capacity cannot be reduced without incurring a more severe potentiometer term penalty and thus limiting feedback, so that the total gain area cannot be profitably increased. The in-band gain can be made greater or less as we concentrate more or less of the total gain area in the transmitted band, but it is found that attempting to get high values of in-band gain area leads to networks which are increasingly sensitive to element deviations. A resistance area efficiency of about 50 per cent turns out to be the most acceptable compromise.

In spite of these steps, the coupling networks are the most important source of manufacturing deviations, but by improved mechanical design and the use of quality control these effects have been reduced by an order of magnitude as compared to previous designs. For example, the end-capacity of each coupling network is a quartz-disc condenser, and the peaking or splitting coils are tension-wound on ceramic forms to give inductances of 1 per cent tolerance with distribution requirements within this range. The windings of the transformers, shown in Fig. 7, are made by plating the turns on threaded forms machined from optical grade quartz or Vycor glass to tolerances of tenths of a thousandth of an inch. Leakage inductance and parasitic capacity deviations are thus held to a minimum. Split ferrite cores, which must also be held to close tolerances, make it possible to use these methods of fabrication, which previously available tape cores would not have permitted.

Since the amplifier configuration gives series feedback on the tubes adjacent to the cable, and cathode feedback on the tubes adjacent to the regulating network, the high impedance winding of each coupling network is necessarily off-ground. This leads to considerable difficulty in specifying the equivalent circuit. For an on-ground coupling network, the equivalent circuit of Fig. 8(b) would be adequate for gain and feedback computations — essentially a reactive equalizer plus an ideal transformer. Even then the capacities and inductances associated with the

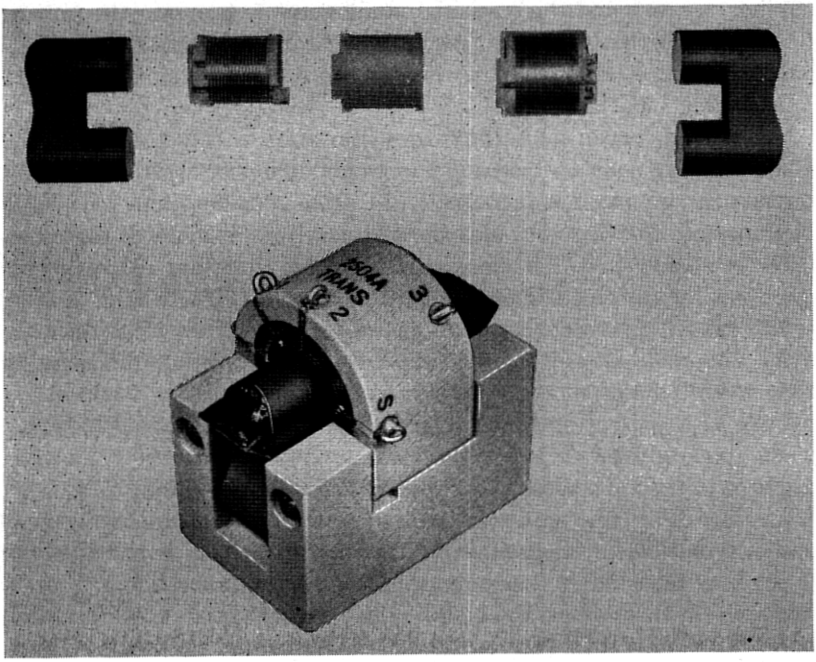


Fig. 7 — Precision transformer, windings, core, assembled unit.

transformer itself would have to be given as functions of frequency because of the distributed nature of the device. When, however, the high impedance winding is raised above ground potential by the voltage developed across the beta circuit, which is nearly the same magnitude as the voltage across the coupling network, the effects of distributed parasitic capacities to ground, and the lumped capacity from the junction of transformer and peaking coil, become of prime importance. The coupling network, therefore, cannot be adequately represented by merely lifting the circuit of Fig. 8(b) off-ground.

In order correctly to understand and compute the amplifier gain, it was necesary to develop a complex mathematical analysis of the distributed structure of the transformer, in conjunction with an extended program of precise measurements of the transformer constants. Even then, one must be content with an accuracy of a few tenths of a db and a few degrees of phase, as compared with the order of magnitude better accuracy which can be obtained for a two-terminal lumped-constant network. The agreement between measurement and computation of amplifier gain and reflection coefficient is sufficiently good, however, to

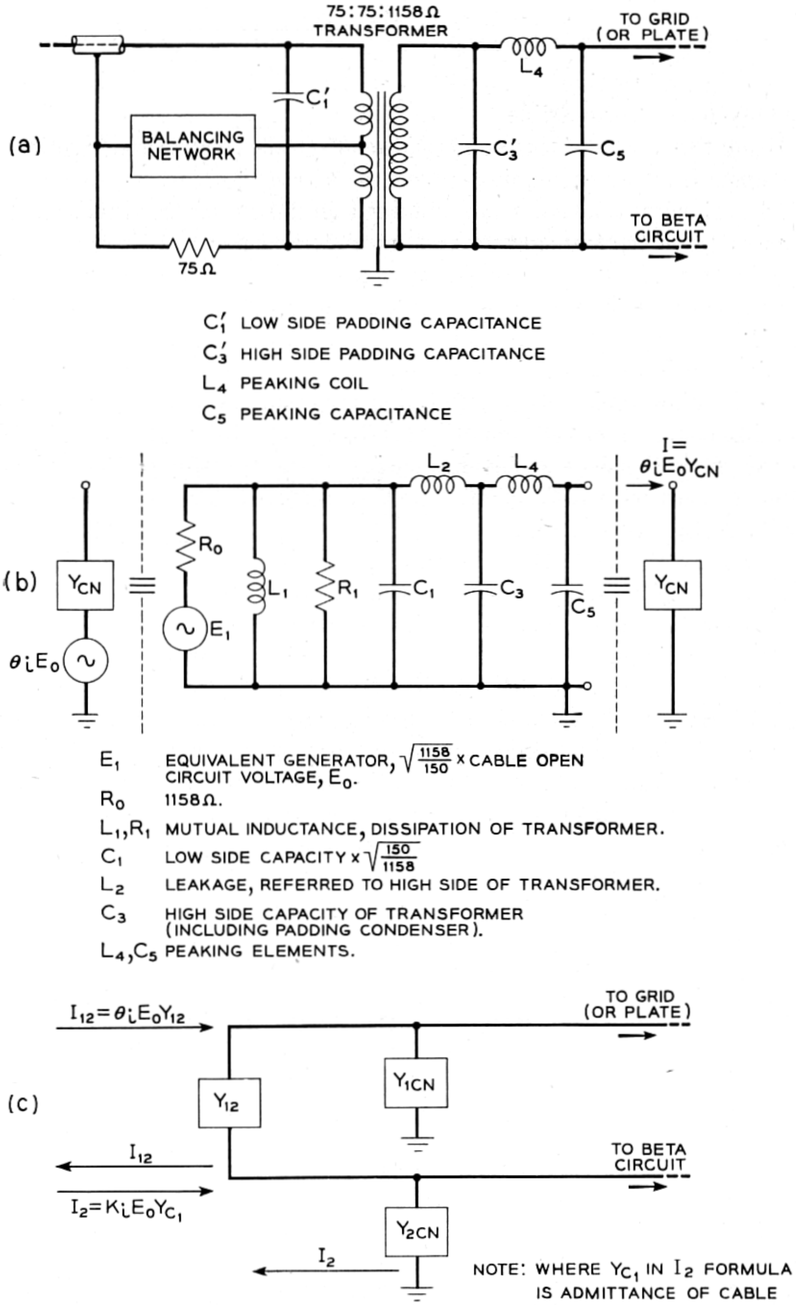


Fig. 8 — Coupling Network Circuits. (a) Physical elements. (b) On-ground equivalent circuit, adequate for gain and feedback computations in an amplifier configuration employing on-ground coupling networks. (c) Off-ground equivalent Circuit.

assure that all the important effects are sufficiently understood to make intelligent control possible.

Using the results of this analysis, the off-ground coupling network can be represented by the equivalent circuit of Fig. 8(c), where the values of the pi of admittances are obtained in terms of the fundamental parameters of the transformer and associated elements. This representation is convenient for insertion gain and feedback computations. The most important difference between the equivalent circuits of Figs. 8(b) and 8(c), from a practical standpoint, is the presence in the latter of the admittance Y_{1CN} which is a manifestation of the capacity to ground from the high impedance winding and the junction of transformer and peaking coil. I_2 and the contributions to Y_{2CN} not directly attributable to the obvious shield-to-shield capacity of the transformer can be set equal to zero with little error, but Y_{1CN} affects the insertion gain and feedback by about 2 db.

It will be observed that the transformer is shown with two shields, one connected to the bottom of the high impedance winding, which is the top of the beta circuit, the other connected to ground. The first of these, which is physically adjacent to the high impedance winding, acts to collect, and carry to the beta circuit, the distributed capacity of the high winding, thus avoiding intolerably large capacity from this winding to ground. The second shield prevents capacitative coupling of the large beta circuit signal voltage to the low impedance winding, which would lead to a very poor reflection coefficient performance. Typical curves of reflection coefficient are shown in Fig. 9. Since the amplifier is an active device, the reflection coefficient is to some extent a function of the vacuum tube transconductances, and tends to be degraded as tubes age.

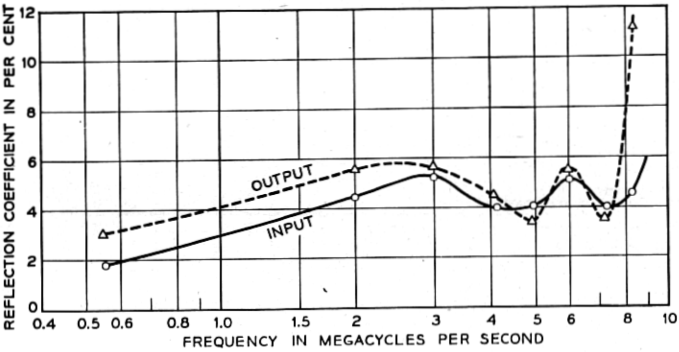
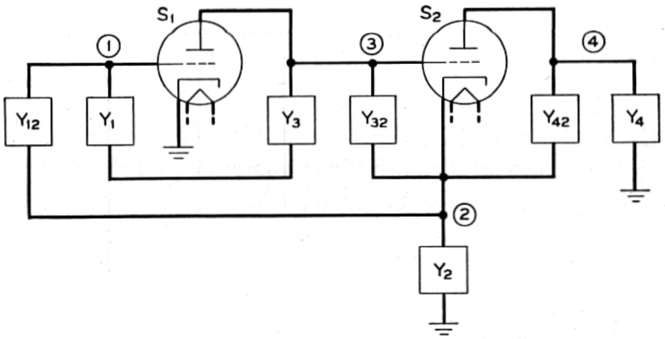


Fig. 9 — Reflection Coefficients.



NODAL DETERMINANT OF CIRCUIT:

$$\Delta = \begin{array}{cccc|c} E_1 & E_2 & E_3 & E_4 & \\ \hline Y_{11} & -Y_{12} & 0 & 0 & I_1 \\ -Y_{12} & Y_{22}+S_2 & -Y_{32}-S_2 & -Y_{42} & I_2 \\ S_1 & -Y_{32} & Y_{33} & 0 & I_3 \\ 0 & -Y_{42}-S_2 & S_2 & Y_4+Y_{42} & I_4 \end{array}$$

WHERE $Y_{11} = Y_1 + Y_{12}$; $Y_{22} = Y_2 + Y_{12} + Y_{32} + Y_{42}$; $Y_{33} = Y_3 + Y_{32}$

FEEDBACKS

RETURN RATIO ON FIRST TUBE: $$T_1 = \frac{P_{12}\dfrac{S_1 S_2}{Y_{33}Y_2'}P_4 + P_{12}\dfrac{S_1}{Y_2'}P_{32}}{1 + P_3\dfrac{S_2}{Y_2'}P_4} = \frac{B+D}{1+E}$$

ON SECOND TUBE: $$T_2 = \frac{P_{12}\dfrac{S_1 S_2}{Y_{33}Y_2'}P_4 + P_3\dfrac{S_2}{Y_2'}P_4}{1 + P_{12}\dfrac{S_1}{Y_2'}P_{32}} = \frac{B+E}{1+D}$$

WHERE $P_{12} = \dfrac{Y_{12}}{Y_{11}}$; $P_4 = \dfrac{Y_4}{Y_4 + Y_{42}}$; $P_3 = \dfrac{Y_3}{Y_{33}}$; $P_{32} = \dfrac{Y_{32}}{Y_{33}}$

$Y_2' = Y_2 + \sigma_1 + \sigma_3 + \sigma_4$; $\sigma_1 = \dfrac{Y_{12}Y_1}{Y_{11}}$; $\sigma_3 = \dfrac{Y_3 Y_{32}}{Y_{33}}$; $\sigma_4 = \dfrac{Y_4 Y_{42}}{Y_4 + Y_{42}}$

TRANSMISSION

$$E_4 = I_{12}\frac{\Delta_{14} - \Delta_{24}}{\Delta} + I_2\frac{\Delta_{24}}{\Delta}$$

USING FOR I_{12} AND I_2 THE VALUES GIVEN ON FIGURE 8(C)

$$E_4 = E_{OC}\left[\theta i\,\frac{Y_2}{Y_4}\,\frac{B - D\dfrac{\sigma_4}{Y_2} - E\dfrac{\sigma_1}{Y_2} - \dfrac{\sigma_1}{Y_2}\,\dfrac{\sigma_4}{Y_2'}}{1+B+D+E} + KiY_{c_1}\,\frac{1}{Y_4}\,\frac{B+E+\dfrac{\sigma_4}{Y_2'}}{1+B+D+E}\right]$$

Fig. 10 (a) — Input amplifier formulas.

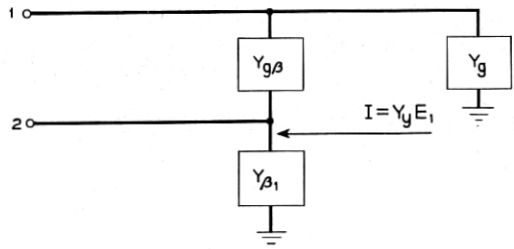
Amplifier Formulas

Using the equivalent circuit of the coupling network shown in Fig. 8(c), the circuit of the input amplifier can be represented as on Fig. 10a, and the formulas shown can be derived from straight forward nodal analysis of the circuit. Similar formulas can be derived for the more complicated output amplifier. From these, using Thevenin's theorem, the gain of the tandem combination can be computed, as well as the feedbacks on the various tubes.

Similarly the input amplifier can be replaced, for convenience in reflection coefficient calculations, by the pi of admittances and the driving current of Fig. 10(b). The formulas of Fig. 10(b) expressed in terms of the co-factors of the circuit determinant are of general application for the reduction of a multi-node circuit to simpler form.

Regulating Network

Like the coupling networks, the regulating network between the amplifiers is outside the feedback loops, and the gain of the amplifier is very nearly a direct function of the impedance seen looking into the network. This impedance is controlled by a single variable resistance — the thermistor — which is directly heated by the dc output current of the regulator. The output of the regulator, in turn, is a function of the



$$Y_g = \frac{\Delta_{22} - \Delta_{21}}{\Delta_{12-12}} = Y_1 \qquad Y_{\beta_1} = \frac{\Delta_{11} - \Delta_{21}}{\Delta_{12-12}} = Y_2 + \sigma_3 + \sigma_4 + P_3 S_2 P_4$$

$$Y_{g\beta} = \frac{\Delta_{21}}{\Delta_{12-12}} = Y_{12} \qquad Y_y = \frac{\Delta_{12} - \Delta_{21}}{\Delta_{12-12}} = -\left[P_{32} S_1 + \frac{S_1 S_2}{Y_{33}} P_4\right]$$

WHERE THE CO-FACTOR Δ_{12-12} IS FOUND BY STRIKING OUT THE FIRST AND SECOND ROW AND FIRST AND SECOND COLUMN OF THE CIRCUIT DETERMINANT, THE SIGN FOLLOWING THE USUAL RULES FOR THE SIGN OF A CO-FACTOR

Fig. 10 (b) — Equivalent circuit of input amplifier as seen by coupling network.

magnitude of main pilot level at the output of the amplifier. The whole forms an AVC circuit which acts to keep the level of the main pilot constant at amplifier output, under the assumption that any changes in this pilot level are caused by line temperature or length deviations. The invariant arms of the regulating network are designed to give the accurate shaping of network impedance versus frequency required so that the amplifier gain change will have the wanted square-root of frequency shape. The design of the network is fundamentally based on the variable equalizer theory developed by H. W. Bode,[5] but the process of finding physical networks which will give the desired performance to a very high order of accuracy makes use of newer methods developed by S.

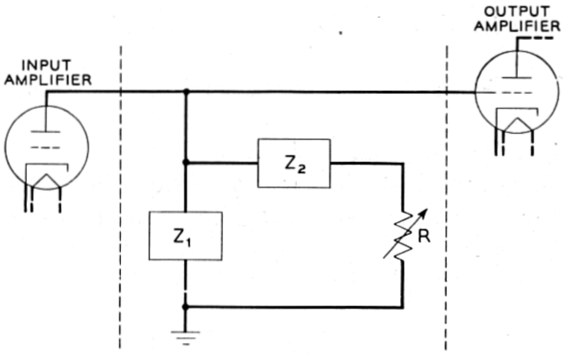


Fig. 11(a) — Regulating network block schematic.

Darlington.[6, 7] Since these methods were also used in the design of several other amplifier networks, and because the precision thus made possible is essential to the system, a brief recapitulation of the steps involved is of interest.

Fig. 11(a) shows the basic configuration of the regulating network. It was selected because it is one of the simplest that gives symmetrical gain changes controlled by a single resistor. It is also capable of absorbing the parasitic interamplifier capacity, and gives some advantageous gain shaping at normal setting, although this shaping is not under design control.

We need to find the impedance of the two arms Z_1 and Z_2, that will give the required square-root of frequency shape of gain change (with a value of 6 db at 8 mc) as the thermistor is varied by a factor of three. For this circuit the change in gain may be expressed as a ratio of two impedances.

$$e^{\alpha+jb} = \frac{Z_H}{Z_N} = \frac{\dfrac{1}{Z_1} + \dfrac{1}{Z_2 + R}}{\dfrac{1}{Z_1} + \dfrac{1}{Z_2 + kR}} = F(p) \tag{1}$$

$$e^{-\alpha-jb} = \frac{Z_L}{Z_N} = \frac{\dfrac{1}{Z_1} + \dfrac{1}{Z_2 + R}}{\dfrac{1}{Z_1} + \dfrac{1}{Z_2 + \dfrac{R}{k}}} = \frac{1}{F(p)} \tag{2}$$

where:

Z_N is the impedance looking into the network with a thermistor value R

Z_H is the impedance looking into the network with a thermistor value kR

Z_L is the impedance looking into the network with a thermistor value R/k

For this design $R = 500$ ohms and $k = 3$

We may solve equations (1) and (2) and obtain expressions for Z_1 and Z_2. However, for synthesis Z_1 and Z_2 should be expressed in terms of singularities in the p-plane. This may be accomplished by approximating the wanted gain change with a Tchebycheff series as developed by S. Darlington. The steps in the process are:

1. The coefficients of a Tchebycheff series that matches the desired gain change over the frequency band of interest are found:

$$\alpha = C_0 + C_2 \cos 2\phi + C_4 \cos 4\phi \cdots C2_n \cos 2n\phi \tag{3}$$

where the relation between ω and ϕ is given by the band-pass transformation

$$\omega^2 = K \frac{k_1 + \sin^2 \phi}{k_2 - \sin^2 \phi}$$

$$K = \omega_{c1}\omega_{c2}$$

$$\omega_{c1} = K \frac{k_1}{k_2} \text{ lower cutoff frequency (.2 mc)}$$

$$\omega_{c2} = K \frac{k_1 + 1}{k_2 - 1} \text{ upper cutoff frequency (8.35 mc)}$$

and $$C_{2k} = \frac{2}{n+1} \sum_{r=1}^{r=n+1} \alpha_r \cos 2k\phi r \qquad \begin{array}{l} k = 1,2....n \\ r = 1,2....n+1 \end{array}$$

(Match points)

$$\phi_r = \frac{(2r - 1)}{(2r + 1)} \frac{\pi}{2}, \qquad \alpha_r = \text{gain at } \phi_r$$

2. The coefficients of a related power series are determined:

$$e^{2\alpha} = 1 + A_2 Z^2 + A_4 Z^4 + \cdots A_{2n} Z^{2n} \equiv F(Z^2) \tag{4}$$

where $\qquad A_2 + S_2 = 0$ and $S_{2k} = kC_{2k}$

$$2A_4 + A_2 S_2 + S_4 = 0$$

$$nA_{2n} + A_{2n-2} + \cdots = 0$$

3. $F(Z^2)$ is expressed as a rational fraction containing both natural modes and infinite loss points.

$$F(Z^2) = \frac{N(Z)^2}{D(Z)^2} \tag{5}$$

where the coefficients of N and D may be found from the continued fraction expansion of $F(Z^2)$. The degree of N and D fixed by the allowable approximation error and the complexity of the network.

4. The roots of N and D (in terms of Z^2) are found and then transformed to the p-plane using the transformation

$$p_\sigma^2 = K - \frac{2K(k_1 + k_2)}{\psi + 2k_2 - 1} \qquad \text{where} \qquad \psi = \frac{1 + Z^4}{2Z^2} \tag{6}$$

These fours steps result in a polynomial $F(p)$ satisfying the requirements on the change in gain. In this specific case

$$F(p) =$$

$$\frac{1.06(p + 0.0960)(p + 0.0280)(p + 0.9890)(p + 0.2890)}{(p + 0.1058)(p + 1.803)(p + 0.0300)(p + 0.3492)} \doteq e^{\alpha + j\beta} \tag{7}$$

Solving (1) and (2) obtain

$$Z_2 = \frac{k - F}{kF - 1} R, \qquad Z_1 = \frac{(k^2 - 1)(F^2 - 1)}{(kF - 1)(k - F)} \tag{8}$$

where $F(p) \equiv F$

Since the design must absorb the interstage capacities, (and also the stray capacities of some of the elements in the physical network) Z_1 must include a shunt condenser. It can be shown that the desired result is obtained when

$$F \to 1 + g \frac{1}{p} \text{ as } \qquad p \to \infty \tag{9}$$

Since the derived transmission function F, equation (7), does not satisfy the condition expressed by (9), the function must be modified in order to provide a shunt condenser in Z_1. To accomplish this we use equation (5) expressed as a bi-linear form of the nth term of the continued fraction expansion.

$$F(Z^2) = \frac{N_1(Z^2) + K_n N_2(Z^2)}{D_1(Z^2) + K_n D_2(N^2)} \tag{10}$$

The coefficient K_n is modified so that the new $F(p)$ satisfies equation (9). The modification of K_n does not substantially alter the required in-band transmission and also does not produce non-physical singularities The new F obtained is

$$F(p) = \frac{(p + 0.0289)(p + 0.1018)(p + 0.3431)(p + 1.017 \pm i\ 1.628)}{(p + 0.0310)(p + 0.1132)(p + 0.4426)(p + 0.6158 \pm i\ 1.368)} \tag{11}$$

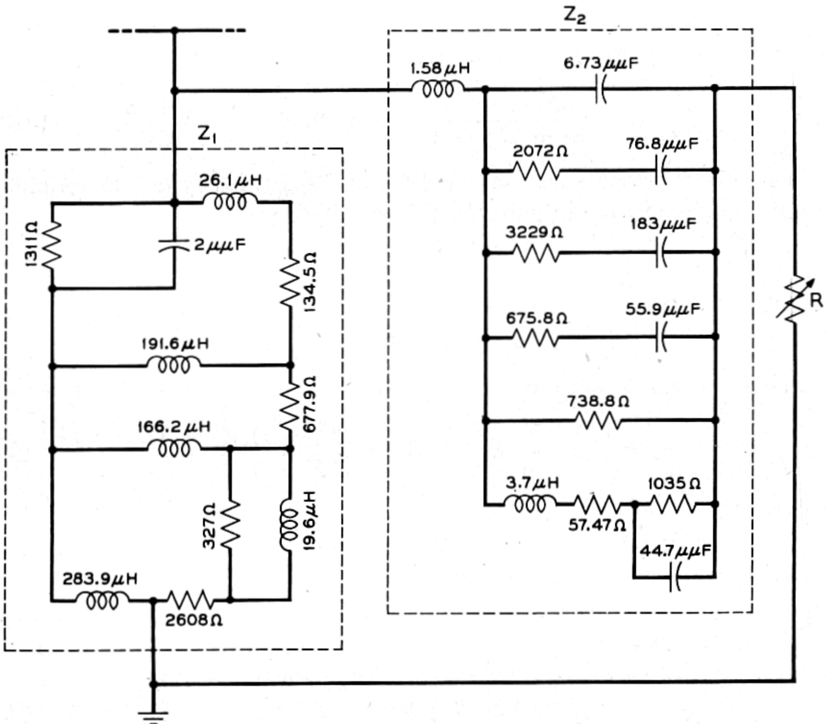


Fig. 11(b) — Regulating network schematic.

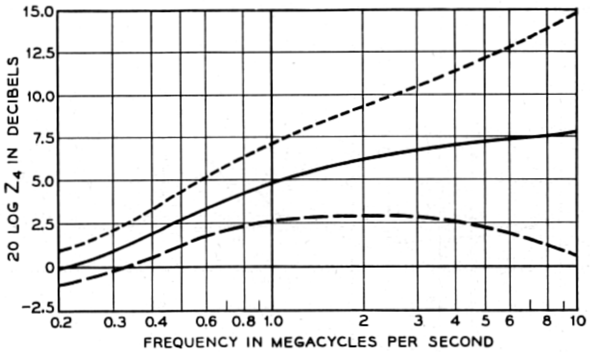


Fig. 12 — Relative gain of regulating network for extreme and mid-range thermistor settings.

By using this result in equation (8) expressions for Z_1 and Z_2 were obtained from which the configuration that is shown in Fig. 11(b) was synthesized.

Fig. 12 shows the voltage which would be developed across the regulating network versus frequency in response to a constant current driving force, for the mid-range and extreme values of thermistor resistance. The slope across the band for the mid-range value of 7.5 db; this is the regulating network contribution to the total slope of 37 db required of the complete amplifier.

To the extent that the differences between the curves of Fig. 12 fail to exactly match the desired square-root of frequency characteristic, the action of the regulating network will introduce an equalization error. This regulation error is shown in Fig. 13. It amounts to a few hundredths of a db for a six db gain change, caused in part by network design imperfections and in part by the fact that very small second order effects result in the amplifier gain not exactly following the regulating network impedance.

Beta Circuits

Starting from our basic requirement that a slope of 37 db across the transmitted band must be obtained, and noting that the total of the contributions from the coupling networks and regulating network is 16 db, we are left with about 20 db of shaping to be supplied by the beta circuits. The input beta circuit has been designed to supply most of this remainder, the contributions of the output beta circuit and the $\mu\beta/(1 - \mu\beta)$ effects in the amplifiers being only three db. The original design procedure for this network was basically the same as for the

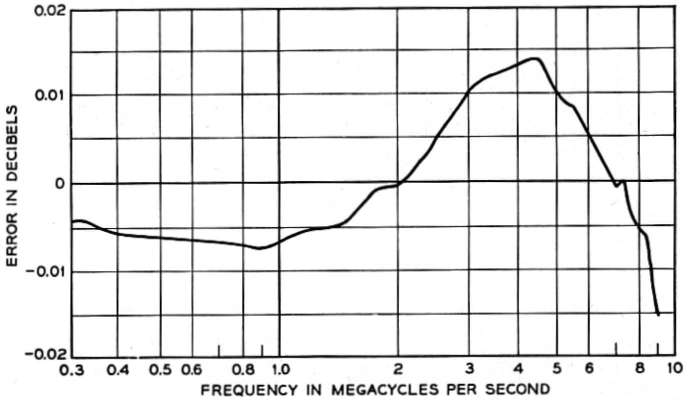


Fig. 13 — Regulation error per one db regulation.

regulating network, but the final values of the elements represent a long process of incorporating secondary corrections as our knowledge of the amplifier grew. Original constraints included the necessity of including as part of the beta circuit not only the relatively large parasitic capacity of the amplifier and the coupling network shield to shield capacity, but also the screen resistor and by-pass condenser of the second tube, which as usual is by-passed not to ground but to cathode, for modulation reasons. It is also required that the beta circuit have the correct dc resistance to serve as the cathode bias resistor of the second tube, and that it incorporate provision for metering this bias through suitable decoupling elements which are among the gain-determining elements of the network. Finally, this network was used as the mop-up equalization network of the amplifier, and its element values were readjusted to give the correct amplifier gain after the performance of a representative group of amplifiers with average coupling networks and tubes had been determined. In doing this tailoring, it is necessary to precorrect for the effects of non-infinite feedback, since the gain of the amplifier is not exactly a direct function of the beta circuit admittance. The configuration of the input beta circuit is shown by Fig. 14; it is a simple two terminal network. Its in-band impedance varies from about 1,000 ohms at 300 kc to 110 ohms at 8.35 mc.

The output beta circuit is relatively flat, which in this case is the optimum condition for signal-to-noise and feedback loop stability considerations. Because of this simplicity, it is possible to incorporate in this network provision for adjusting the gain of the amplifier to reduce the misalignment of the system.

MISALIGNMENT ADJUSTMENT

We can distinguish three major effects which will contribute to misalignment and consequent degradation of system signal-to-noise ratio. One is design error — the degree to which the design gain of the amplifier, because of the finite number of elements and other limitations, fails to match the line loss. Second is the cumulative effect in the individual amplifier of manufacturing deviations of the elements. Third is the aging of the components of the amplifier, of which the tube aging will of course be the dominant short term effect. The signal-to-noise performance of the system can be improved by reducing the misalignment, if this is done without resorting to measures which would introduce systematic instead of random gain deviations.

If we study the shapes of gain change introduced by the more important deviation contributors, particularly the element variations, we find as might be expected that in general the effects are small at low frequencies and increase sharply near the upper edge of the band if the thermistor is held constant. In the system, the regulation around the main pilot will automatically act to reduce the deviation at 7 mc to zero, adding a square root of frequency curve that results in a bow shaped deviation as illustrated by Fig. 15. Examining the signal-to-noise effects of the degradation caused by misalignment, we conclude that it is most desirable to reduce as much as possible the misalignment at 4 mc, the television carrier frequency in the combined system. The output beta circuit has, therefore, been designed to give varying amounts of this shape, the total range being ±0.6 db at 4 mc in 0.2 db steps. The successive steps of this gain adjustment are simple multiples of each other, symmetrical in the two directions of adjustment, so that we put into

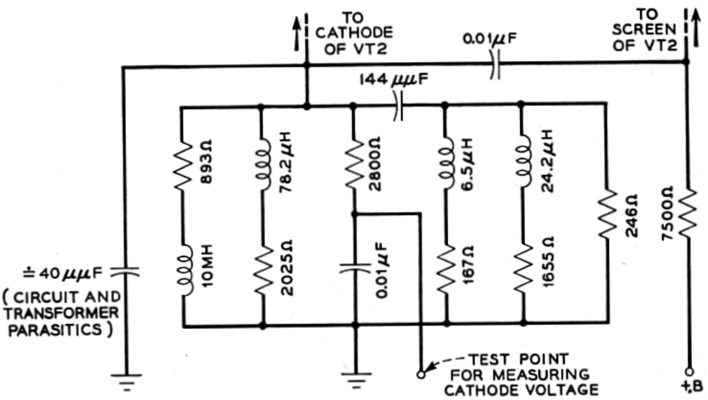


Fig. 14 — Input beta circuit elements — line amplifier.

the system a single systematic shape, which will be small in magnitude at the equalizing points because the different settings of this control in successive line repeaters will tend to cancel out. This setting of output beta circuit gain is controlled by a GAIN ADJ switch accessible from outside the amplifier housing. The adjustment will be made on an in-service basis as part of normal maintenance procedures, using the level at each repeater of one of the system pilot frequencies (3,096 kc) as the index of proper setting.

Manufacturing Testing

As mentioned above, the mechanical design of the amplifier has been planned to permit the separate testing of the five gain-determining net-

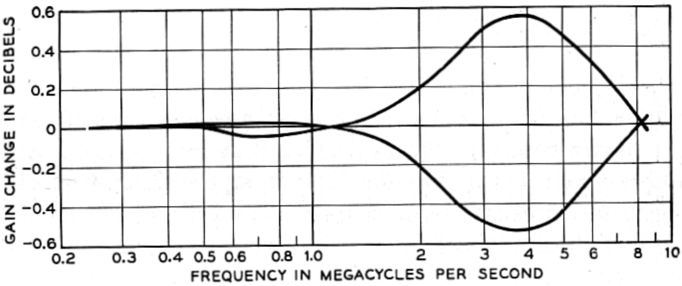


Fig. 15 — Amplifier gain versus output beta circuit misalignment adjustment.

works, the tuned interstage of the input amplifier, and the separate input and output amplifiers before their assembly with the regulating network to form complete amplifiers. Variations in environment are minimized by the use of jigs which also make it unnecessary, in general, to solder to the network under test. Visual gain sets which cover the transmitted band and are accurate to two or three hundredths db are used. The component network or component amplifier under test is connected in series with a complementary or equalizing network to obtain a flat transmission characteristic which can be accurately compared to the transmission of standard attenuators. The completely assembled amplifiers are similarly tested. Quality control charts of the resulting measurements on all components are useful in detecting shifts in transmission which might be caused by loss of control in element manufacture or by shifts of element values caused by subsequent damage in handling.

DC Feedback

In addition to the loop feedback at signal frequencies, local dc feedback is used on each stage. The grid of each tube is returned to a +9-volt potential rather than to ground, and about +10 volts is developed across each cathode resistor. The usual stabilizing effects of self-bias are thus obtained in exaggerated degree, each tube having about 20 db of local dc feedback. Care must then be taken to select the cathode by-pass and interstage coupling condensers so that the transition from low-frequency local feedback to in-band loop feedback is accomplished smoothly without the instability which might be caused by a balancing out of these two feedbacks in the transition region.

For maintenance measurements, the 9-volt bias potential and the dc cathode voltage of each tube are brought out to a multi-pin amplifier test jack through appropriate decoupling filters. Thus the bias on each stage can be measured on an in-service basis. Filament voltage dropping resistors which can be switched in or out of circuit are provided on the repeater panel, so that bias can also be observed for a filament supply voltage 10 per cent below normal. The activity or change in plate current with filament voltage thus measured, or the history of the bias at normal filament voltage, will be used to determine when amplifiers should be taken out of service for tube replacement.

The upper triode is, of course, a special case: since it is the plate supply path for the lower triode, its cathode is about 160 volts above ground and its grid is returned to a similarly high potential. About 35 db of dc feedback is obtained on this stage; the same provisions for measurement of bias and activity are made. To avoid excessive filament to cathode voltage, a separate filament winding which floats at the +190-volt supply potential is used for this tube.

Loop Feedback

The design of the feedback loops of the input amplifier follows conventional practice. The constraints which operate to limit the amount of feedback which can be obtained in the transmitted band are well known.[8] Broadly speaking, the figure of merit of the vacuum tubes and the circuit capacities determine the asymototic cutoff, which in any feedback amplifier limits the magnitude of the feedback which can be built up in the band. In multi-loop structures, there are additional limitations. Consider, for example, the formula given on Fig. 10(a) for the feedback on the second tube. If we increase without limit the magnitude of the first tube transconductance, we find that the feedback on the second tube approaches the value $S_1 P_4 Z_{32}$. This is a limit, common to multi-loop structures, over and above the usual limit imposed by the Nyquist stability criterion. A similar $S\ P\ Z$ limit can be derived for the feedback on each stage. The same multi-loop mechanisms which operate to limit the feedback also result in making the feedback on any given stage relatively insensitive to variations in other transconductances or impedances. In a single loop circuit a one db change in beta circuit impedance or first stage transconductance causes a one db change in the feedback on every tube. Here the feedback on, say, the second stage would generally change only one-half db at most frequencies. While this is an advantage in the sense that transconductance decay does not decrease feedback as rapidly as it would in a single-loop amplifier, it is a disadvantage in that it militates against improving stability margins by shaping the out-band impedance of the beta circuit.

The capacity distributions within the input and output amplifiers are such that while the $S\ P\ Z$ limitations on feedback are closely approached, the most stringent limitation on the feedback obtained is the Nyquist criterion.

The use of the Nyquist criterion, particularly with respect to defining the margins against singing, is likewise complicated by the multi-loop nature of the circuit. Ignoring some very recent work, the implications of which have not been fully explored at this writing, it can be said of a multi-loop structure that the apparent margins against singing shown by any plot of feedback give no certain information as to how safe from singing the circuit is. The phase, as well as the magnitude, of the feedback on each stage is a function of the magnitude of the other transconductances, and either decay or increase of these transconductances might destroy the phase margin. In these circumstances, it is theoretically necessary to examine for stability every conceivable combination of transconductances. A more practical expedient, of course, is to rely on judgment backed by computation and laboratory experiment on the circuit for a wide but far from infinite number of circuit conditions.

Another difficulty arises from the fact that the feedback on each tube is different, so that gain and phase margins obtained for one return ratio do not imply equal gain and phase margins for the return ratio on some other tube of the same amplifier. It does not follow, however, that it is necessary to investigate separately the margins on each stage versus circuit element variations. The point to be stressed here is that we are using the behaviour of the return ratio merely as an index of our real concern: the position on the p-plane of zeros of the determinant of the circuit, and the determinant is the same for both stages. Rather than asking how many db of gain margin and how many degrees of phase margin any particular return ratio displays, we ought instead to inquire how quickly the apparent margins disappear as we change transconductances and network impedances. The margins on all the return ratios will vanish simultaneously, regardless of the apparent difference in original margin magnitudes. It is therefore satisfactory to examine the behaviour of whichever return ratio is most easily observed.

The design choices made in arriving at the coupling network also affect the magnitude and shape of the feedback which can be obtained: as mentioned above, the relative magnitude of the impedance seen looking into the coupling network and the impedance from first tube grid to ground determines how much of the voltage developed across the beta circuit will reach the first stage as a driving force.

Study of the modulation products which will arise in system operation, both for the all telephone case and for the combined telephone-television signal, led to the conclusion that optimum shaping of the feedback for the L3 system would be to maximize the feedback at low frequencies in order to suppress intermodulation products falling in that part of the spectrum in the combined telephone-television case. Shaped feedback, falling off at the higher frequencies, is also consistent with obtaining a smooth and simple shape of gain-change as tubes age (known as "mu-beta effect"), which is desirable from the equalization standpoint. With these considerations in mind, the interstage of the input amplifier has been peaked well above the transmitted band — at 11 mc — to partially compensate for the input potentiometer term, and to help in achieving this smooth shape of mu-beta effect. If flat feedback over the band were the objective, it would also be necessary to design the grid-cathode admittance of the second tube so that the parasitic grid-cathode capacity would be absorbed in a flat impedance versus frequency, but in this case the desired shaping of the second tube feedback is attained by taking advantage of the way in which the grid-cathode capacity naturally limits the high-frequency feedback on this stage.

The loop feedback in the output amplifier is similarly shaped, for the same modulation and equalization reasons. The use of the double triode circuit in this amplifier is an unusual feature. This connection of two triodes, sometimes referred to as the "cascode circuit," has appeared in many contexts in recent years, usually to serve some other purpose than here. It serves as a superior output stage in the L3 amplifier largely because the effective transconductance which can be obtained is about 3 db higher than that of a pentode of the same grid-cathode spacing. The effective transconductance, ignoring for the moment the division of output current between the load impedance and the internal plate impedance of the upper triode, is very nearly the transconductance of the lower triode. Since no current is lost to a screen, this is higher than that of a pentode of the same spacing. The output impedance of the upper triode is multiplied by the local feedback consequent on its cathode being off ground by the plate impedance of the lower triode. Since the load presented to the lower triode is a low impedance, approaching the reciprocal of the transconductance of the upper triode, the grid-plate capacity of the lower triode is not enhanced by feedback. The higher value of transconductance means less grid drive for the same output current, and more obtainable feedback, both contributing to superior modulation performance.

The modulation of the upper triode, and the changes in transconductance of this tube, are suppressed by approximately the sum, in db, of the local and loop feedback. In consequence, the modulation and mu-beta effect contributions of this tube to the complete amplifier are negligibly small.

It will be observed that the grid of the upper triode is not connected to ground, but to the top of the beta circuit. In consequence, the grid-plate capacity of the upper triode appears as part of the end-capacity of the coupling network rather than as a parasitic capacity to ground. This results in a gentle potentiometer term in the output amplifier. This connection, however, also has the effect of vitiating to some extent the desirable qualities of the circuit, particularly at the higher frequencies, where the grid-cathode capacity of the upper triode becomes important.

Because of this gentle output potentiometer term, it is not necessary to tune the interstage of the output amplifier, which consists simply of the circuit capacity plus a network which has the characteristics of a 10-mmf capacity in the transmitted band. Above the band this network shapes the gain and phase characteristics of the feedback loop to obtain the desired stability margins. The incorporation of this network reduces the in-band feedback by about 3 db, a sacrifice which unfortunately is necesary to assure stability when the thermistor in the regulating network is at its minimum value. For this value of thermistor, the phase and magnitude relations of the regulating network impedance and the grid to cathode capacity of $VT3$ produce a potentiometer term in the feedback loop which appreciably reduces the margins around 30 mc. The stability margins for the mid-range value of thermistor would be satisfactory without this sacrifice of in-band feedback. When the thermistor is at maximum resistance, some degradation of phase margin at 10 mc occurs, again because of the potentiometer term effect mentioned above, but in this case the remaining margin is sufficient, since the circuit elements are still under good control at this frequency. Because the plate-cathode impedance of $VT2$ is very high, the similar potentiometer term at the output of the input amplifier causes only negligible changes in input amplifier stability margins as the thermistor changes.

In the 70-mc region there are two almost equally important feedback loops in the output amplifier — one through the transconductance of the lower triode, the other through the grid-plate capacity of this tube. Balances between these feedback paths are observed in the 70 to 100-mc region in the course of measuring the feedback, sometimes accompanied by 180° shifts in the phase of the loop transmission at frequencies above the balance point, an effect which theoretically depends on just how the two vectors go through the balance point. The occurrence of these balances is accompanied by a few degrees loss of phase margin in the 30-mc cut-off region, which must also be allowed for in setting the 30-mc stability margins, since sufficient control of parasitics to prevent these 70-mc effects is out of the question.

Parasitic resonances between the lead inductances and the capacities of the circuit, which tend to cause instabilities in the very high-frequency region about 200 mc, are damped by small resistors in the leads, and the lead inductances are kept small by careful mechanical design. In this frequency region, neither measurement nor computation of stability margins can be trusted as anything but a rough guide. On the other hand, adding damping resistors in grid leads and other critical points to prevent 200-mc sings causes a phase margin penalty in the 30-mc region, so a nice judgment of how much damping to add is called for. Final values of damping were chosen so that typical amplifiers could not be made to sing by increasing critical lead lengths or by substantial increases in parasitic capacity, thus assuring that manufacturing variations of elements and wiring will not cause high-frequency sings. The return ratio of $VT4$ is shown on Fig. 16, the mu-beta effects of both amplifiers on Fig. 17.

Signal Levels, Modulation, and Noise

Fig. 18 shows the signal levels within the amplifier in db relative to one volt from grid to cathode of $VT4$, which is a convenient point to use as a reference for system signal-to-noise studies. It will be noted that as a result of using the input beta circuit to give so much of the shaping of amplifier gain, the input amplifier has little gain at low frequencies. In consequence the input tube of the output amplifier and the regulating network are important thermal noise sources at low frequencies. The

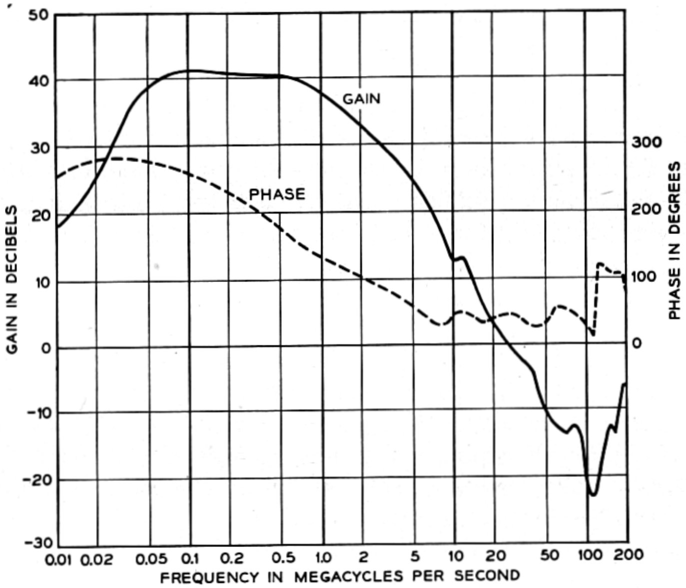


Fig. 16 — T4, return ratio of lower triode.

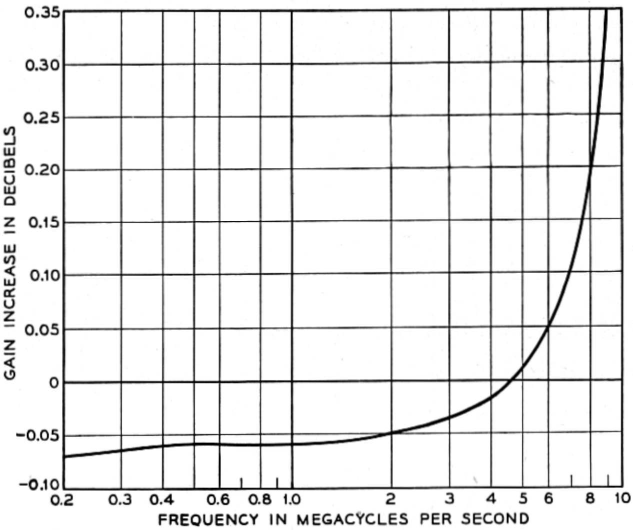


Fig. 17 — Mu beta effect, amplifier gain change for a one db decrease in each transconductance.

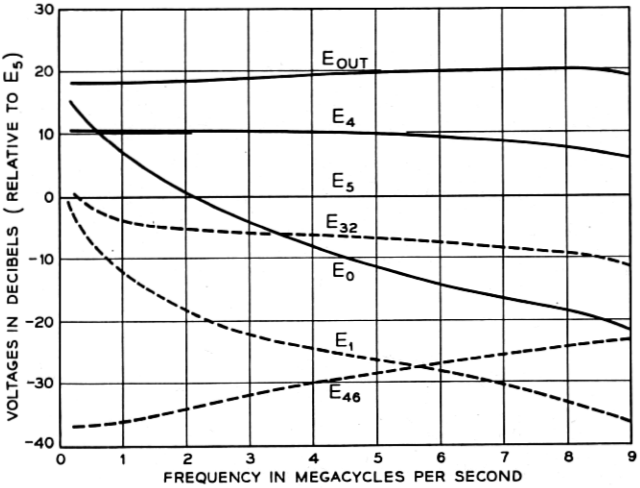


Fig. 18 — Relative levels of signal voltages in line amplifier referred to voltage at grid of lower triode. Thermistor at mid-range value.

relative magnitudes of the noise sources are shown on Fig. 19, which gives the noise at amplifier output as a function of frequency.

Comparison of the grid to cathode voltages of $VT2$ and $VT4$ shows that the former will be an important modulation contributor, since the driving force on these tubes is nearly equal, particularly at low frequencies. Typical amplifier modulation values are given in Table III. Computations using the measured feedback and the performance of

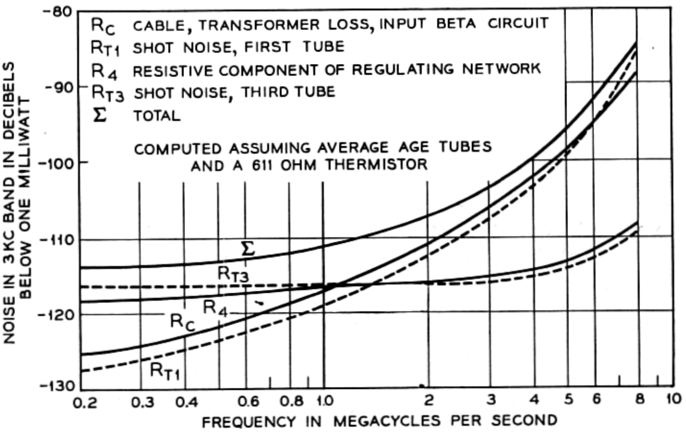


Fig. 19 — Thermal noise at line amplifier output.

single tubes without feedback check the measured values of amplifier modulation to within a couple of db if the third order coefficient of the tube is corrected to take account of the fact that some third order modulation is generated by the interaction of the fundamentals and the fed-back second order products. In general, the effective third order coefficient of the tubes is approximately equal to the voltage sum of the tube's uncorrected third order coefficient and a coefficient 6 db worse than the square of the tube's second order coefficient. If this interaction correction is not taken into account, the correlation of tube modulation, feedback and amplifier modulation is unsatisfactory. The analysis leading

TABLE III — MODULATION PRODUCTS, IN db BELOW ONE MILLIWATT AT AMPLIFIER OUTPUT, FOR FUNDAMENTALS 5 db ABOVE ONE MILLIWATT AT AMPLIFIER OUTPUT

Type	Fundamentals Mc	Product Mc	Product −dbm
2F	0.5	1.0	72
	1.0	2.0	65
	3.0	6.0	55
	4.0	8.0	51.5
3F	0.25	0.75	95
	0.667	2.0	97
	2.0	6.0	87.5
	2.66	8.0	81.5

to this result, which is due to F. B. Llewellyn, S. E. Miller and R. W. Ketchledge, is too long to give here.

LOAD CARRYING CAPACITY

The load carrying capacity of an amplifier is difficult to define with exactness. One possible definition is the load at which the modulation coefficients of the amplifier have departed appreciably from the small signal power series values because of loss of feedback as the transconductance is cut off during part of the cycle. The signal carried without serious overload, in terms of a single frequency, is practically constant in the transmitted band as a consequence of the fact that the output voltage and the lower triode grid voltage have nearly the same shape versus frequency, as shown on Fig. 18. The output coupling network shaping approximately compensates for the potentiometer term division of current between the load impedance and parasitic paths to ground. Departure from the small signal power series behaviour just begins to be appreciable at +14 dbm of any single frequency. At +18 dbm the dc effects of overload show up as slight changes in the transmission of the pilot frequencies. At +26 dbm, the second order modulation is 3 db, the third order modulation is 6 db, higher per line amplifier than would be predicted from small signal behaviour.

Flat Amplifier

The flat gain amplifier, which is used as a transmitting amplifier and to make up for equalizer loss at various points in the system, is basically the same as the line amplifier, with only the obviously necessary modifications. The input beta circuit is nearly flat, the regulating network has been replaced by a fixed gain network which contains a single variable element whose adjustment at the factory compensates to some extent for variations in the coupling networks, and the input coupling network has been modified so that the peaking used in the line amplifier is replaced by a drop in the high-frequency gain of this network. The output beta circuit has been modified to give flat gain control of ±1.0 db in 0.2 db steps. The interstage designs are changed to readjust the feedback so that the modulation suppression and the change in gain as tubes age will be nearly the same as in the line amplifier. Somewhat more feedback is obtained in the output amplifier since no network is needed in the output amplifier interstage to adjust the 30-mc phase for an unfavorable regulating network setting.

The nominal gain of the flat gain amplifier has been set at 34 db and is flat to within ±0.2 db over the transmitted band. The amplifier circuit capacities are low enough so that the inter-amplifier network could be built to give considerably more gain than this; the limit has been set so that flat gain amplifier noise contributions to the complete system noise will not exceed about 1.0 db at the television carrier.

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References

1. C. H. Elmendorf, A. J. Grossman, R. D. Ehrbar, R. H. Klie, The L3 Coaxial System — System Design, see pp. 781 to 832 of this issue.
2. R. W. Ketchledge, T. R. Finch, The L3 Coaxial System — Equalization and Regulation, see pp. 833 to 878 of this issue.
3. G. T. Ford and E. J. Walsh, Electron Tubes for a Coaxial System, Bell Sys. Tech. J., Oct. 1951, Part II.
4. H. W. Bode, Coupling Networks, U. S. Patent 2337965, Dec. 28, 1943.
5. H. W. Bode, Variable Equalizer, Bell Sys. Tech. J., April, 1938.
6. Sidney Darlington, The Potential Analogue Method of Network Synthesis, Bell Sys. Tech. J., April, 1951.
7. Sidney Darlington, Network Synthesis Using Tchebycheff Polynomial Series; Bell Sys. Tech. J., July, 1952.
8. Hendrik W. Bode, Network Analysis and Feedback Amplifier Design, D. Van Nostrand, 1945.