The Transistorized A5 Channel Bank for Broadband Systems

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This article presents a brief historical background to the A-series of terminal units, used extensively in long-haul and short-haul transmission facilities to provide the first step of modulation from voice to carrier and the final step of demodulation from carrier to voice. Most of the paper is devoted to a description of the latest unit of this series — the A5 channel bank, which through the use of transistors and other modern components achieves significant improvements in size, power requirements and operating characteristics.

I. INTRODUCTION

All of the long-haul, broadband transmission facilities of the Bell System, and many of the short-haul microwave radio systems, employ a common unit in their terminal multiplexes. This is the A type channel bank which provides the first step of modulation from voice to carrier spectrum and the reverse function of demodulation. From its inception over twenty-five years ago, the channel bank has undergone a number of size reductions primarily utilizing new types of crystal filters. This paper describes a radically new version, the A5 Channel Bank, with a detailed discussion of the new circuit feature — the transistorized voice-frequency amplifier.

II. HISTORICAL BACKGROUND

2.1 *General*

Over many years, the threads of the "channel bank" story have been woven into papers of much broader scope. It is the aim of this section to combine these pieces into one background picture.

In the middle 1930's, Bell System research and development effort in long-distance communication was focusing on the expansion of frequencydivision, single-sideband carrier systems.[1] The invention of the negative-feedback amplifier by H. S. Black had made feasible the development of transmission media capable of carrying multi-channel, high-quality systems across the country.[2] Other important advances in both the electronic art and the network art opened the way to the realization of practical multiplexes to translate many voice or other information channels to spectra suitable for line transmission.

At this time, many important decisions were made on multiplexing methods. Since that time, they have set the broad pattern for the terminals of long-haul carrier systems, both for wire and microwave radio. Inclusion in the recommendations of such bodies as the CCITT* made the general pattern worldwide. The decision of greatest interest to this paper involves the first step of modulation from voice frequency to a carrier spectrum and its counterpart of demodulation at the receiving end.

Actually, not one, but many considerations were involved. The fundamental ones concerned carrier spacing, the number of voice channels to be handled in the modulation process, and the frequency spectrum to be used. The answers did not evolve without much thought directed toward the future long-distance plant and the carrier systems which would make it a reality. The development trend was toward the utilization of much higher frequencies for line transmission than had previously been attempted.[3] Three high-frequency broadband systems were being conceived for application to:

1. Available 19-gauge toll cable, both underground and aerial (Type K).[4,5]
2. Open-wire lines then carrying the three-channel Type C system. The new system (Type J) would lie above Type C in frequency and require extensive retranspositions of the line.[6]
3. A new medium — coaxial cable — with much greater potential for high-frequency transmission. This would be the transmission medium for the L1 and L3 systems of the future.[7,8,9,10]

As the study progressed, it became more and more obvious that, if possible, the same terminal arrangements should serve all of these systems.

2.2 *Carrier Spacing*

Of primary importance in the pattern for the new systems was the question of carrier spacing. Assuming that the new broadband transmis-

* Comité Consultatif International Téléphonique et Télégraphique.

sion media did not have sharp cut-offs, the penalty for wide carrier spacing was more repeaters per link. Of course, the open-wire and cable systems did have practical limitations such as crosstalk, which presented economic but not insurmountable restrictions. Contrariwise, the new coaxial line seemed restricted only by attenuation due to thermal noise and by repeater spacing. Under these circumstances, the cost penalty per voice circuit due to wide spacing would not be controlling. For short systems where terminal costs predominate, the cost was insignificant; even for very long systems, it was quite small.[11]

The standardization of 4000-cycle spacing for broadband systems was based primarily on other considerations. The Bell System was striving to improve the over-all quality of the service it was offering. Subscriber sets, telephone instruments, and end links were all being developed to provide better quality. It seemed logical to make the backbone carrier circuits good enough not to be limiting in the over-all effective net loss. The 4000-cycle spacing made this possible. With advances in the filter art, the width of the frequency band could be extended at both low and high ends. The former helped to maintain the naturalness of the voice; the latter helped to improve articulation. At this time it was also decided to make the carrier intervals exactly uniform — in other words, to relate all carriers harmonically to 4 kc. This was not a matter of quality of voice transmission, but one of terminal equipment.

A corollary of high-capacity broadband systems is heavy concentrations of equipment at major terminal offices. Under these circumstances, supplying of carrier power poses an economic problem. Needed are many frequencies of high precision both initially and with time and changing environment. Economic feasibility of producing many harmonically related frequencies of approximately equal amplitudes had been demonstrated.[12] For the new broadband systems the circuit took the form of a highly stable 4-kc source driving a pulse producer. Odd harmonics were directly generated; even harmonics were provided by rectification. Sufficient power was derived to feed many systems from common bus-bar arrangements.

2.3 *The Modulation Process*

Of prime importance in devising a multiplexing plan is the decision as to single- or multi-stage modulation from voice to line frequency. Systems prior to the broadband development employed a single stage. For the new systems, several factors influenced the choice of multi-stage modulation.

It was known early that the line frequencies of the three systems — cable, open-wire and coaxial — would be quite different. Economic restrictions rather than technical feasibility imposed by crosstalk and line equalization set a practical limit for the long-distance, cable-pair system between 10 and 60 kc. Since the new open-wire system was to be placed above the standard Type "C", the lower limit was about 35 kc; the upper limit of about 150 kc was dictated by the costs of line transpositions. The coaxial cable system seemed to have a reasonable lower limit of about 60 kc, but an upper limit in the order of one megacycle. The latter was soon to become three megacycles, and finally, for L3, about eight megacycles. For the open-wire and cable-pair systems, single-stage modulation was possible but the terminals, except for an overlap region between 30 and 60 kc, required different channel filters and carrier supplies. Single-stage modulation was out of the question for the high channel capacity coaxial system. It would mean hundreds of different, closely spaced channel filters not yet feasible even with the rapidly advancing network art. Also, for every channel a different carrier frequency would be needed with resultant terminal complications.

The answer to these problems was the provision of a group of channels common to all three systems. With 4-kc spacing, both the cable-pair and open-wire systems could economically accommodate 12 channels. Of course, the coaxial system did not have this restriction, but 12 channels seemed a good common denominator.

This resolved the technical differences; in addition, the choice of a common group would mean (1) flexibility of interconnection of systems, (2) large-scale production of one major equipment unit, and (3) minimum development effort in system areas and in the supporting network and component areas as well.

2.4 *The Group Spectrum*

The foregoing discussion on the size of the group might seem to imply a completely free choice. Perhaps, at the present time, it could be made independently of the spectrum to be employed. This was not true in the 1930's. The limitations of available inductor-capacitor filters in the low-frequency ranges made a 12-channel group impracticable, even starting at a low frequency such as 12 kc. The first experimental cable system had demonstrated the difficulties above about 40 kc. It had provided nine channels with this top frequency.

Of course, as was done later in certain European and U. S. systems, another stage of modulation could have been introduced and the subgroup technique used, i.e., three or four channels for instance, operating in the most efficient low-frequency range translated to a 12-channel group.

However, the expanding art had provided another answer. Intensive research and development in the Laboratories by Mason, Sykes, and Lane had culminated in successful application of the piezoelectric effects in quartz crystals to wave filters.[13,14,15] These were the same effects which had been studied by Langevin and applied to submarine detection in World War I and by Cady, Pierce, and others.

The decision was made to employ such crystal filters for the selection of the desired single sideband. They offered many advantages. The transmission bands could be positioned in the frequency spectrum with high precision. Steeply rising attenuation characteristics, low distortion and low flat loss could now be achieved rather easily.

The actual choice of 60 to 108 kc as the standard group frequency range of the channel bank was mainly based on economics. Lower frequency crystals were physically possible. They were large and expensive and, if high production was envisioned, might be difficult to obtain from the raw Brazilian quartz available. Crystals at frequencies higher than 108 kc were small and easily available, but their fabrication and frequency adjustment posed difficult problems in the 1930's. Thus, the final frequency choice of 60 to 108 kc as the group spectrum to provide 12 channels was an economic and technical compromise.

2.5 *Microwave Radio Usage*

As described, the channel bank was developed in the 1930's, long before the potentialities of the microwave radio band were realized. However, when radio systems at frequencies in the 4-kmc range (TD-2 Radio) were being developed, their high channel capacities could utilize the available coaxial system terminals based on the channel banks.[16,17] Interconnection with wire systems was thus made easy, and the high production needs of the new radio systems further proved the value of a standard and basic group of channels. Later microwave radio systems, TH and TJ, also employ the standard "A" type bank.[18,19]

III. OBJECTIVES OF A5 DESIGN

From these beginnings, the channel bank has progressed through several previous redesigns. Aimed specifically at size reduction, they were based mainly on advances in the crystal filter art. These had led to smaller filters as well as to much improved crystal arrangements.[20]

The A5 design had broader objectives. Modernization and miniaturization were important, but equally so were transmission improvement and maintenance simplification. The introduction of new services such as Direct Distance Dialing and data transmission has imposed new requirements on the long-distance plant. Specifically, the circuits must maintain better net loss stability. This must be true in the face of office voltage variations and changes in the performance characteristics of the active devices employed.

With the ever-expanding, long-haul toll plant, the maintaining of performance within close limits becomes a problem of large proportions. Improved circuit designs and devices can in the long run avail little if proper maintenance is made difficult. The A5 bank design aimed at facilitating this important feature of Bell System service. Accessibility of adjustment and ease of interchanging parts were important factors in the design. Circuit compatibility with existing long-haul terminal gear was another important requirement.

IV. GENERAL CIRCUIT FEATURES

For the channel banks, the general circuit arrangements have remained unchanged through the several equipment redesigns, including the A5. The channel bank equipment does not include circuits for signaling or for the conversion from two-wire to four-wire operation. This permits greater flexibility in circuit arrangements. As shown in Fig. 1, the transmitting circuit starts at the voice-frequency input transformer, whose leakage reactance also offers high impedance to carrier frequencies. A simple shunt varistor bridge employs copper oxide as the modulating element. The modulator operates at a nominal carrier power level of 0 dbm. The high-pass filter following the modulator provides high impedance to voice frequencies, thus increasing the modulator efficiency. The desired lower sideband is then selected by a crystal band-pass filter operating in parallel with eleven others. A compensating network in parallel with the common output improves the transmission characteristic of the channel filters. The hybrid output coil provides an alternate output to facilitate switching a working bank to an alternate group facility without interruption. It also provides a means of inserting a program terminal occupying the frequency space of two or three regular voice channels.

The receiving circuit operates in much the same fashion as the transmitting circuit except, of course, in reverse. The high-frequency line terminates in a hybrid transformer, which provides for connection to a program receiving terminal. The twelve channel filters are paralleled at this hybrid. The individual sidebands are then demodulated in a bridge

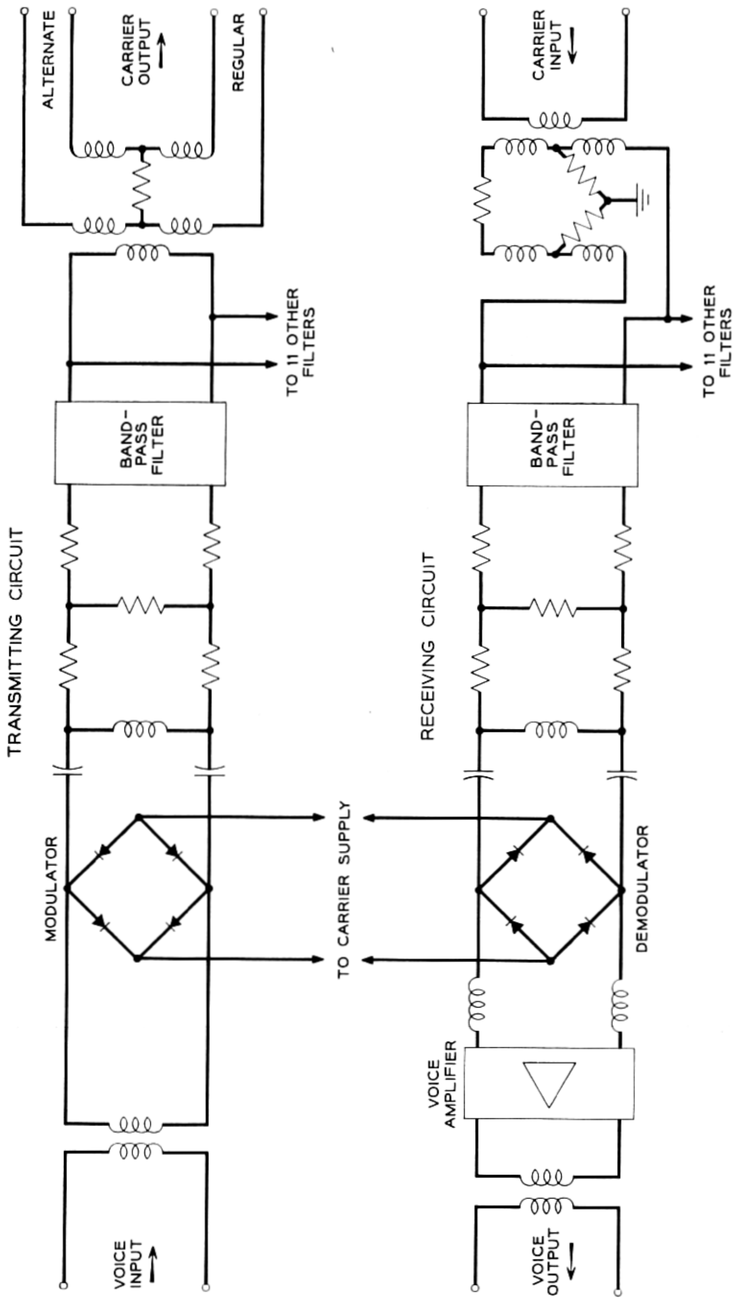

Fig. 1 — Basic circuit of the A5 channel bank.

similar to the modulator. The demodulator, however, is poled oppositely on the carrier supply to neutralize the dc components of the modulation process in the two units. High-pass and low-pass filters terminate the demodulator to improve its efficiency.

Following the low-pass filter is a voice amplifier with sufficient gain to establish the correct received voice level. The potentiometer for gain control is remotely mounted at the four-wire patching bay in the test area of the toll office.

From their inception, as described previously, the Bell System channel banks have been based on a single modulation process from voice to carrier output and upon the crystal filter as the sideband selecting medium. Other methods are feasible and have been used in this country as well as in Europe. Pregrouping both at low and high frequencies has been used, with the standard 60- to 108-kc group spectrum being obtained by a second stage of modulation. With the availability of high efficiency ferrite inductors, it is also possible to obtain good LC filter performance, usually with the addition of some band equalization. Economic studies indicated that, with Western Electric's highly developed capabilities in crystal filter production, the single modulation step using such filters was still the least expensive procedure.

Design choices had to be made regarding two other major circuit units — the modulator-demodulators and the only active unit for each channel, the voice amplifier at the output of the demodulator. In previous banks the modulating unit, a quad of copper-oxide varistors, had proved generally satisfactory. During the A5 development, newer devices such as silicon diodes were studied. However, problems of incompatibility with the carrier power available at existing offices and other considerations made it advisable to employ the standard copper-oxide varistors.

In previous banks, the demodulator amplifier employed a single electron tube and provided approximately 35 db of gain in each channel over the voice range. Only modest feedback was used with a maximum value of 7 db at minimum gain. This was reduced to zero at maximum gain. With the small amount of feedback, the tube amplifier was sensitive to tube aging, battery variations, and component changes. The resultant variations in net loss were tolerable for the voice plant with switching under operator control. The introduction of Direct Distance Dialing and the need to transmit various data signals impose more severe requirements. It is difficult to meet these consistently without frequent and very careful maintenance.

An important part of the A5 development was the improvement of this demodulator amplifier. The main focus, of course, was on the provision of sufficient feedback to overcome the instabilities of the existing design. Two design approaches were available, a multi-tube or a transistor amplifier. With well-established techniques, the former offered no particular design problems; however, it did mean heavier power consumption, more heat to be dissipated, and a definite restriction on the degree of miniaturization possible.

On the other hand, a transistor amplifier offered a reduction in power, a cooler operating environment, and the possibility of a high degree of size reduction. In addition, the promise of very long life could mean a large reduction in maintenance cost. On the basis of these considerations, the decision was made to proceed with a transistor channel bank.

V. TRANSISTOR AMPLIFIER

The design specifications for the amplifier were:

1. Voltage gain between 600-ohm terminations should be adjustable between 30 to 40 db.

2. The voltage gain should have a 5-db rise at 200 cycles in order to compensate for the low-frequency rolloff of the crystal band-pass filters in the modulator and demodulator.

3. A particular gain adjustment should not change by more than ± 0.1 db over a period of years and for ± 10 per cent variations in the -24-volt central office battery.

4. Amplifier should provide 80 milliwatts of output power into a 600-ohm load with less than 1 per cent second and third harmonic distortion.

5. Output noise power should be less than 1.4×10^{-7} milliwatts (6.5 dba at zero-level point).

6. Return loss at input and output circuits should be greater than 40 db over the frequency band of 200 to 4000 cycles.

In addition to the above requirements, it is desirable that the amplifier be as simple as possible and employ the minimum number of components.

5.1 *Basic Design Considerations*

In order to satisfy the requirement that the gain of the amplifier be constant over a relatively long period of time, it is evident that a feedback amplifier is required. A hybrid feedback or a series feedback connection at the output of the amplifier appears to be most attractive. Hybrid feedback has the advantage over other types of feedback in permitting a 50 per cent reduction in dc power dissipated in the output stage. Series feedback, on the other hand, has the advantage that for a given amount of negative feedback, it yields a somewhat better return loss than hybrid feedback. In addition, the series feedback circuit is simpler than the hybrid circuit and permits the use of a less expensive output transformer. For these reasons, it was decided to use a series feedback connection at the output of the amplifier rather than hybrid feedback. Similar considerations of return loss and circuit simplicity applied to the input of the amplifier indicate that series feedback is again the optimum connection.

It is demonstrated in Appendix A that the voltage gain of the series feedback amplifier shown in Fig. 2 is given by the expression

$$G_V = \frac{R_L}{Z_{12F}} \cdot \frac{A\beta}{1 - A\beta} \tag{1}$$

where R_L is the load resistance, Z_{12F} is the open-circuit transfer impedance of the feedback network ($R1$, $R2$ and $Z3$) and $A\beta$ is the loop current transmission. (In the case of transistor feedback amplifiers, it is convenient to define feedback as a loop current transmission.[21]) It is evident from (1) that if the magnitude of $A\beta$ is much greater than one, the voltage gain is determined by the load resistance and the feedback network. In order to achieve the desired voltage gain and distortion performance, R_L is of the order of several hundred ohms, and the magnitude of the transfer impedance $|Z_{12F}|$ is of the order of several ohms. Therefore, $R1$, $R2$ and $Z3$ are relatively small impedances, and $R2$ (a potentiometer) can be mounted at a considerable distance from the amplifier without introducing excessive electrostatic pickup.

It is shown in Appendix A that the input and output impedances of the series feedback amplifier are equal to

$$Z_{\mathrm{IN}} = (Z_{\mathrm{IN}}' + Z_{11F})\,[1 - A\beta(R_G = 0)] \tag{2}$$

$$Z_{\mathrm{OUT}} = (Z_{\mathrm{OUT}}' + Z_{22F})\,[1 - A\beta(R_L = 0)] \tag{3}$$

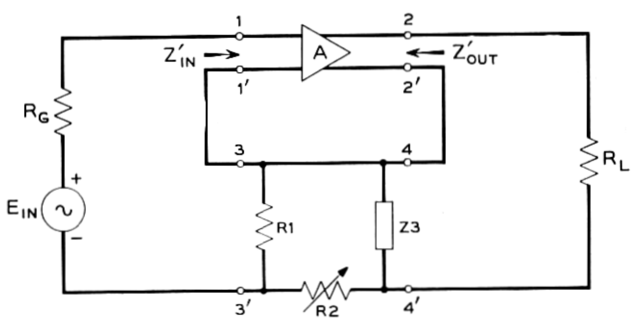

Fig. 2 — Equivalent circuit of a series feedback amplifier.

where Z_{IN}' and Z_{OUT}' are the input and output impedances respectively of the amplifier A (refer to Fig. 2) with the feedback loop opened, Z_{11F} and Z_{22F} are the open-circuit input and output impedances respectively of the feedback network, and $A\beta(R_G = 0)$ and $A\beta(R_L = 0)$ are the values of loop current transmission with R_G and R_L respectively set equal to zero. Since the magnitude of the loop current transmission is much greater than one and Z_{IN}' and Z_{OUT}' are of the order of 1000 ohms, a good return loss can be obtained by shunting the input and output circuits of the feedback amplifier with 600-ohm terminating resistors.

In order to satisfy electrical requirements 3, 4 and 6 in the above list, a minimum of 30 db of negative feedback is required. This amount of feedback and 40 db of external voltage gain can be obtained with three common-emitter connected junction transistors in the amplifier A. If two transistors are used in the amplifier, one of the transistors would have to be connected in the common-collector configuration and the other in the common-emitter configuration in order to provide the phase reversal necessary for negative feedback. Under these conditions the design would be marginal, and severe electrical requirements would have to be placed on the alphas of the transistors (alphas in excess of 0.995 would be required).

Fig. 3 shows a simplified circuit diagram of the amplifier. The dc biasing and the feedback shaping networks have been omitted. Strictly speaking, this structure is not a "pure" series feedback amplifier because the emitter of the transistor in the second stage and the interstage networks are returned to the common connection between the input and output transformers instead of to the emitters of the input and output

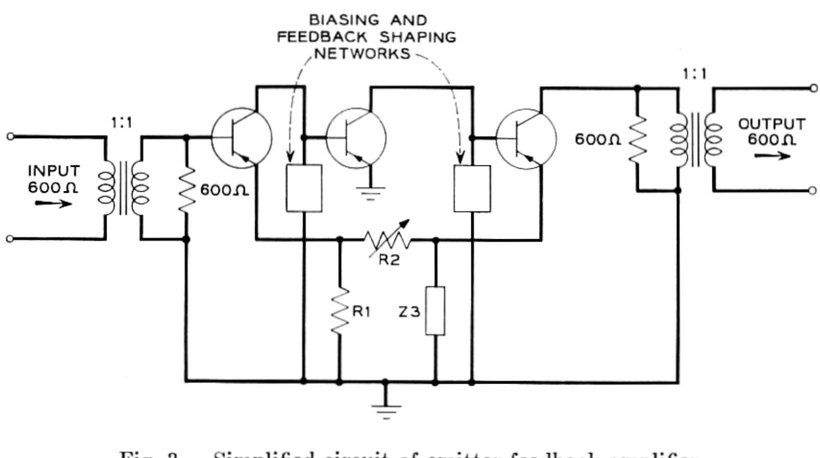

Fig. 3 — Simplified circuit of emitter feedback amplifier.

transistors. This type of feedback structure is defined as an emitter feedback amplifier and is analogous to the cathode feedback circuit for electron tubes.[22] In the case of the series feedback circuit, the network $R1$, $R2$, and $Z3$ provides feedback only around the main loop. In the case of emitter feedback, the network provides local feedback for the first and third stages of the amplifier in addition to feedback around the main loop.

The principal advantage of emitter feedback over series feedback, in a transistor amplifier, is that emitter feedback often makes possible a simpler biasing circuit. Emitter feedback, however, has two disadvantages. First, since it introduces local feedback into the first and third stages, it increases the alpha requirements on the transistors in order to obtain the necessary amount of loop feedback. This disadvantage is partially compensated for by the fact that the local feedback improves the return loss at the input of the amplifier and the distortion performance of the output stage. Secondly, emitter feedback stabilizes the emitter current of the output stage instead of the collector current (refer to Appendix A). As a result, the expression for voltage gain (1) must be multiplied by the alpha of the output transistor. This is not a serious limitation in the case of the demodulator amplifier since the gain of the amplifier is initially set by the use of a potentiometer. In addition, it is expected that alpha will not vary by more than ±2 per cent over the life of a transistor.

5.2 *Magnitude of Feedback in the Useful Operating Band*

In order to calculate the feedback, it is convenient to redraw the circuit as a series feedback amplifier, as shown in Fig. 4. The local feedback

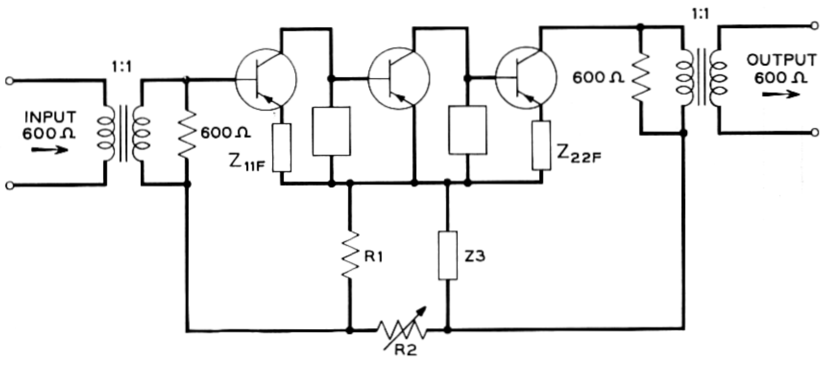

Fig. 4 — Series feedback amplifier equivalent to emitter feedback amplifier.

introduced by the feedback network is taken into account (to a good approximation) by the impedances Z_{11F} and Z_{22F} in the emitters of the first and third transistors respectively. It is evident from (2) and (3) that if the magnitude of $A\beta$ is sufficiently large, the input and output impedances of the amplifier will be equal to 600 ohms, and the load resistance, R_L, is equal to 300 ohms. Since the voltage gain of the amplifier must be adjustable from 30 to 40 db, and allowing for a 2-db loss in the input and output transformers, the transfer resistance of the feedback network must be adjustable between 7 and 2.2 ohms as determined by (1) for $R_L = 300$ ohms. The open-circuit transfer impedance of the feedback network is equal to

$$Z_{12F} = \frac{R1 \cdot Z3}{R1 + R2 + Z3}. \tag{4}$$

The resistance $R2$ is used to control the flat gain of the amplifier while the impedance $Z3$ controls the shape of the gain-frequency characteristic which is required to have a 5-db rise at 200 cycles. With reference to (1) and (4) it is evident that if $|Z3|$ is much less than $(R1 + R2)$, then the voltage gain of the amplifier is proportional to $1/|Z3|$. An additional requirement on $R1$ and $|Z3|$ is that they be as small as possible in order to minimize the amount of loop feedback lost through local feedback. $R1$ was chosen as 68.1 ohms while the flat-gain value of $|Z3|$ was chosen as 10 ohms. From (4), the maximum value of $R2$ (corresponding to maximum gain) is equal to 270 ohms.

In Appendix A it is shown that the loop current transmission of the amplifier shown in Fig. 4 is equal to

$$A\beta = \frac{-G_i Z_{\text{OUT}}' Z_{12F}}{(Z_{\text{IN}}' + Z_{11F} + R_G)(Z_{\text{OUT}}' + Z_{22F} + R_L) - Z_{12F}^{\,2}} \tag{5}$$

where G_i is the short-circuit current gain of the amplifier A. In the frequency range of 200 to 4000 cycles per second,

$$G_i = \frac{a_{01}}{1 - a_{01}} \cdot \frac{a_{02}}{1 - a_{02}} \cdot \frac{1}{1 - a_{03}} \qquad (6)^*$$

$$Z_{\text{IN}}' = r_{b1}' + \frac{r_{e1} + Z_{11F}}{1 - a_{01}} \tag{7}$$

$$Z_{\text{OUT}}' = r_{c3}(1 - a_{03}) \tag{8}$$

* With reference to Fig. 3, the feedback network is driven by the emitter current of the output stage, and as a result this transistor acts as a common collector stage as far as the feedback is concerned. Therefore, the current gain of the third stage is equal to $1/1 - a_{03}$ instead of $a_{03}/1 - a_{03}$ as determined from the equivalent series feedback amplifier shown in Fig. 4.

where a_{01}, a_{02}, and a_{03} are the low-frequency common-base current gains (alphas) of the first, second, and third transistor stages respectively, r_{e1} and r_{b1}' are the emitter and base resistances respectively of the transistor in the first stage and r_{c3} is the collector resistance of the output stage. Substituting (6) to (8) into (5) yields the expression for mid-band feedback:

$$A\beta_0 = \frac{-\dfrac{a_{01}a_{02}Z_{12F}}{1 - a_{02}}}{[(r_{b1}' + Z_{11F} + R_G)(1 - a_{01}) + r_{e1} + Z_{11F}] \cdot \left[1 - a_{03} + \dfrac{R_L + Z_{22F}}{r_{c3}}\right] - \dfrac{Z_{12F}^2(1 - a_{01})}{r_{c3}}}. \tag{9}$$

It will be evident that for all practical designs, the second term in the denominator of (9) can be neglected. As previously discussed, a minimum of 30 db of negative feedback is required in order to satisfy the electrical requirements. The loop feedback is a minimum when the amplifier is adjusted for its maximum gain ($| Z_{12F} | = 2.2$ ohms) and the transistors used have the minimum allowed values of alpha.

$$\begin{aligned} a_{01} &= a_{02} = a_{03} = 0.975 \\ r_{b1}' &= 100 \text{ ohms*} \\ r_{e1} &= 13 \text{ ohms} \\ r_{c3} &= 50{,}000 \text{ ohms.} \end{aligned}$$

If the above transistor parameters are substituted into (9), the resulting magnitude of $A\beta_0$ is 30 db. Consequently, the minimum magnitude of feedback is about 30 db over the frequency range of 400 to 4000 cycles per second. Due to the increase in voltage gain at frequencies between 200 and 400 cycles per second, the minimum magnitude of feedback in this frequency range can be less than 30 db.

5.3 *DC Biasing and Low-Frequency Shaping of the Negative Feedback Amplifier*

Fig. 5 shows a complete circuit diagram of the transistor amplifier. The first stage is biased at two milliamperes of collector current. This value of current is a compromise between low noise figure (requiring a low collector current) and a large magnitude of feedback [requiring a large

* The values stated for the transistor parameters r_{b1}', r_{e1} and r_{c3} are average values for the particular transistors used in the amplifier. Refer to Section 5.3 below for a discussion of the transistors used and the dc operating points.

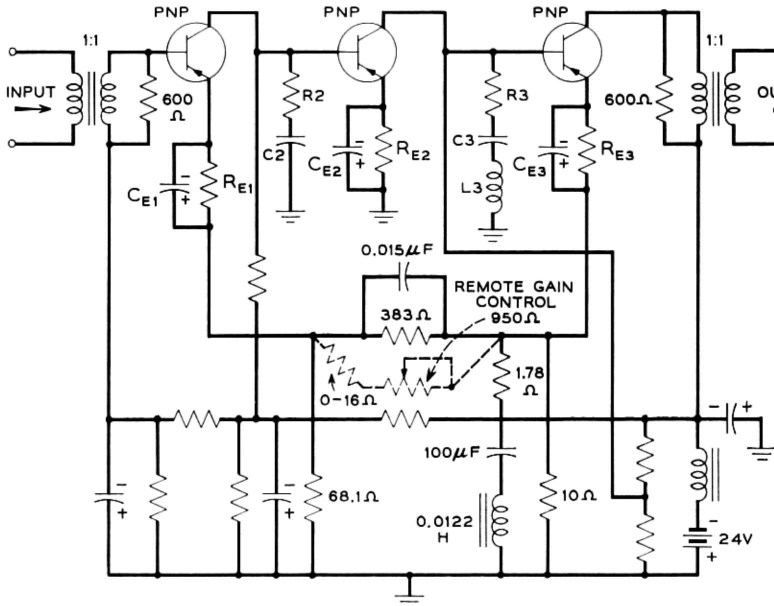

Fig. 5 — Complete circuit diagram of the transistor amplifier.

collector current for low emitter resistance — refer to (9)]. The second transistor stage is biased at 4.7 milliamperes of collector current while the third stage is biased at 45 milliamperes of collector current in order to satisfy the output power requirement of 80 milliwatts into a 600-ohm load. The first and second stages are biased at 2.5 volts collector-to-emitter voltage, while the third stage is biased at 14 volts collector-to-emitter voltage. The collector currents are stabilized by use of dc feedback provided by resistors in the emitter circuits of the transistors.

Type 12B alloy germanium PNP transistors are used in the first and second stages of the amplifier, and a type 9D alloy germanium PNP transistor is used in the output stage. The 12B is a low-power device limited to about 100 milliwatts of dc power dissipation at a 60°C ambient, while the 9D is a medium-power transistor capable of handling 1.5 watts of dc power with an adequate heat sink at a 60°C ambient. Since the 12B transistors in the first and second stages are biased at 5 milliwatts and 12 milliwatts of dc power respectively, and the 9D is biased at 630 milliwatts of dc power, good transistor reliability should be realized.

One of the most important considerations in the design of a feedback amplifier is shaping the negative feedback at high and low frequencies

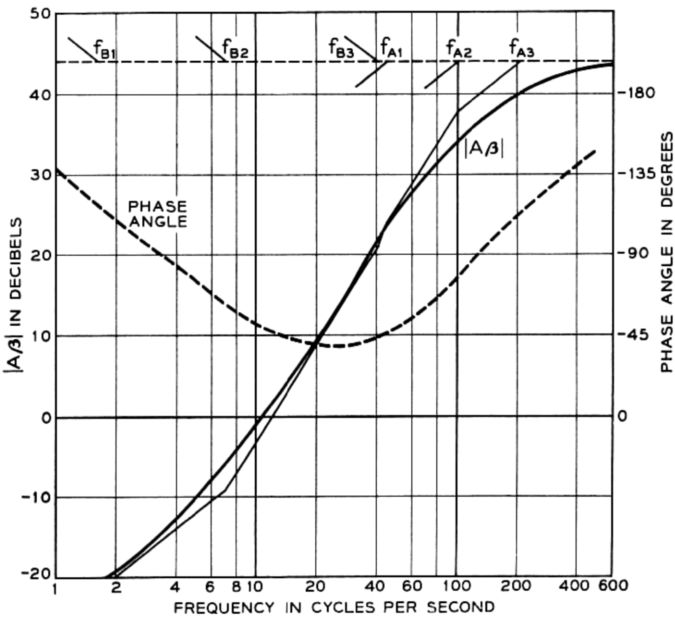

Fig. 6 — Plot of low-frequency loop current transmission with transistors as characterized in Table I.

in order to insure stability for all transistors that will be used in the amplifier. A detailed analysis of the low-frequency shaping of the negative feedback is presented in Appendix B. Fig. 6* shows a plot of the low-frequency loop current transmission calculated for transistors with the electrical parameters listed in Table I and for maximum negative feedback ($| Z_{12F} | = 7$ ohms). At the frequency 25 cycles, the phase of the feedback makes its closest approach to the critical phase angle of 0 degrees. It is evident that the amplifier has a low-frequency phase margin of 40 degrees. If transistors with alphas of 0.99 are used in the amplifier, the phase margin is reduced to 25 degrees. Since the phase of the feedback does not reach 0 degrees, it is not possible to define a gain margin.

5.4 *High Frequency Shaping of the Negative Feedback*

A detailed analysis of the high-frequency feedback shaping is presented in Appendix C. The results of that analysis will be used to show

* In this figure it is assumed that $Z3$ is equal to its flat gain value of 10 ohms. The variation of $Z3$ with frequency has negligible effect on the low-frequency stability.

TABLE I

$a_{01} = a_{02} = a_{03} = 0.975$
$f_{a1} = f_{a2} = f_{a3} = 3$ megacycles
$c_{c1} = c_{c2} = c_{c3} = 35\ \mu\mu f$
$r_{b1} = r_{b2} = r_{b3} = 100$ ohms
$r_{e1} = 13$ ohms
$r_{e2} = 5.5$ ohms
$r_{e3} = 0.6$ ohms
$r_{c3} = 50{,}000$ ohms
$m_1 = m_2 = m_3 = 0.2$ radian

that the amplifier is stable at high frequencies for all transistors for which

$$C_c \leqq 35\ \mu\mu f\ (|V_{cE}| = 2.5 \text{ volts}) \tag{10}$$

$$f_a \geqq 3 \text{ megacycles}. \tag{11}$$

Fig. 7 shows a plot of the high-frequency loop current transmission calculated for transistors with electrical parameters listed in Table I and for maximum feedback. At the frequency 133 kc, the magnitude of the

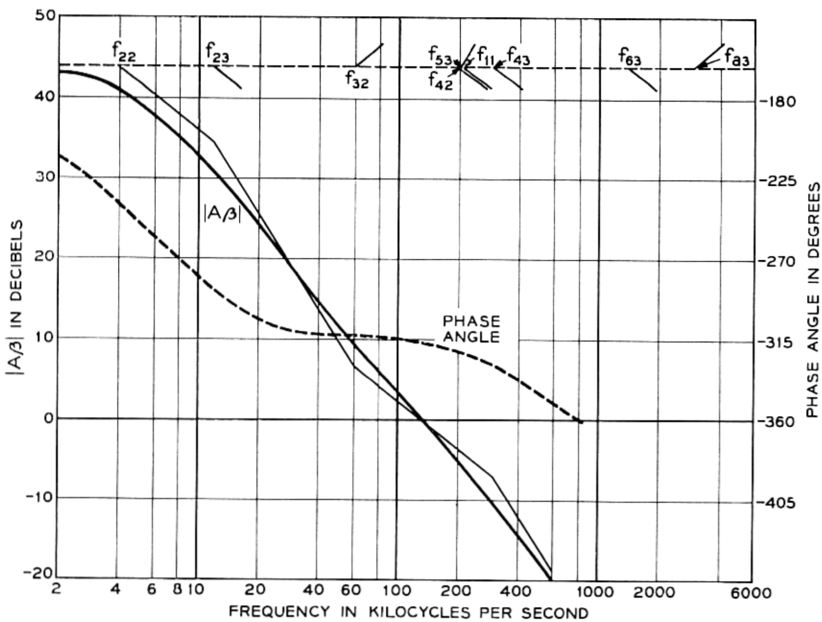

Fig. 7 — Plot of high-frequency loop current transmission with transistors as characterized in Table I.

feedback is 0 db while the phase of the feedback is -317 degrees. At the frequency 800 kc, the phase of the feedback is -360 degrees while the magnitude is -24 db. The phase and gain margins against instability are 43 degrees and 24 db respectively.

Except for the first two cutoff frequencies f_{22} and f_{23}, all of the other cutoff frequencies are essentially independent of the low-frequency common-emitter current gain $(a_0/(1 - a_0))$ and are determined by the transistor parameters f_a, r_b', and C_c and the circuit elements in the two interstage shaping networks, $R2$, $C2$, and $R3$, $C3$, $L3$. The cutoff frequency f_{22} is equal to the frequency at which the reactance of the capacitor $C2$ is equal to the input resistance of the second transistor stage, while the cutoff frequency f_{23} is equal to the frequency at which the reactance of $C3$ is equal to the input resistance of the third stage. To a good approximation, the input impedance of a common emitter stage is directly proportional to the current gain of the stage.* Since the current gain of a common-emitter stage may vary from 39 to 200 (corresponding to alpha variations of 0.975 to 0.995), the cutoff frequencies f_{22} and f_{23} may be as small as one-fifth the values shown in Fig. 7. Fortunately, this variation in cutoff frequencies f_{22} and f_{23} is almost exactly compensated for by the variation in current gain of the second and third stages respectively, and the asymptotic loop current gain is independent of the common emitter current gain at frequencies above f_{23}. The first stage of the amplifier acts as a common-base stage as far as the feedback is concerned. With reference to (9), it is evident that the magnitude of the feedback is essentially independent of the factor $(1 - a_{01})$ if $|(r_{e1} + Z_{11F})|$ is much greater than $|(r_{b1}' + R_G + Z_{11F})|(1 - a_{01})$. These results are important since it means that the high-frequency stability of the amplifier is essentially independent of the low-frequency common emitter current gain.

If transistors are used that have high-frequency parameters superior to those listed in Table I, then the high-frequency stability of the amplifier is improved. In particular, an increase in alpha cutoff frequency or a reduction in collector capacitance or base resistance will tend to increase cutoff frequencies f_{11}, f_{42}, and f_{43}, thus providing larger gain and phase margins against instability.

5.5 *Electrical Performance and Gain Control Circuit*

The amplifier satisfies all of the electrical requirements (1) to (6). Fig. 8 shows a plot of the closed loop voltage gain of the amplifier. To

* This approximation assumes that the component of input resistance due to emitter resistance is much larger than r_b'. This approximation is valid for the transistors in the amplifier.

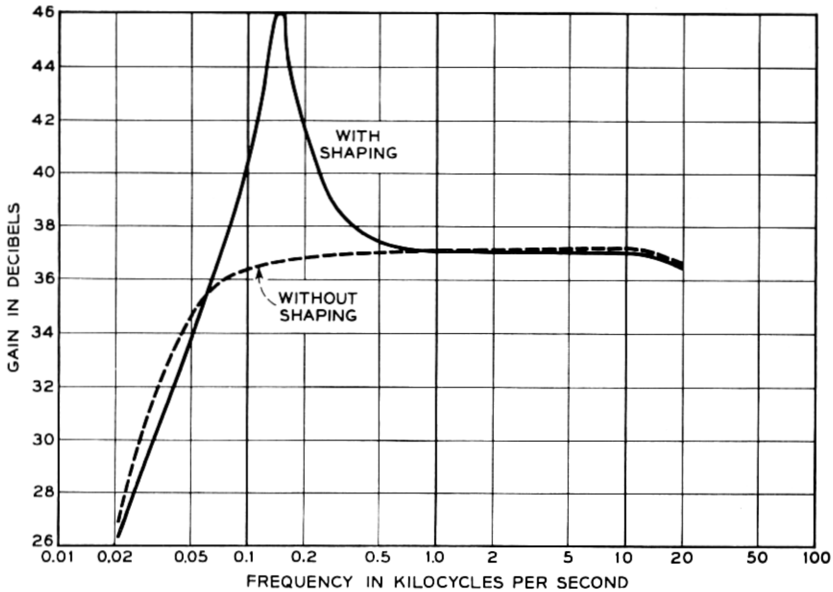

Fig. 8 — Closed loop voltage gain of the amplifier.

a good approximation, the shape of the gain-frequency characteristic is independent of the flat gain.

The gain control circuit consists of a 950-ohm potentiometer in series with the resistance of a 22-gauge wire with a length that may vary between 0 and 1000 feet. The resistance of the connecting wire has a maximum value of 16 ohms which can be neglected. In Section 5.2 it was pointed out that $R2$ had a maximum value of 270 ohms. This is obtained by placing a 383-ohm resistance across the remote gain control circuit as shown in Fig. 5. A 0.015-microfarad capacitor is also placed across the remote gain control circuit in order to minimize the effect of undesirable impedance variations introduced by the connecting wire acting as a transmission line at high frequencies.

VI. EQUIPMENT DESIGN

The equipment features of the A5 bank differ greatly from those of any previous long-haul carrier arrangements. In achieving the new format the chief objectives were: (1) miniaturization, (2) economical manufacture, (3) economical installation, and (4) simplified maintenance.

Taking advantage of the transistor amplifier and newly available passive components such as ferrite transformers, a substantial size reduction is achieved. The new bank, which is now a single equipment unit rather than a grouping of panels, is designed for 19-inch rack mounting. It occupies $12\frac{1}{2}$ inches of vertical height and has an over-all depth of 10 inches. Fig. 9 shows an A5 unit compared to an A4 bank. With presently available fuse arrangements and inter-bay cables, nine A5 banks can be provided on an 11-foot 6-inch bay. This represents a 3:1 improvement in space utilization over the A4 bank. It is expected that with development now in progress on new fuses and on smaller cables, ten A5 banks will be mounted in the 11-foot 6-inch bay. A picture of the size reductions from the first designs until now is given in Fig. 10. A close-up front view of the A5 bank is shown on Fig. 11. An A5 bank weighs about 70 pounds as compared to 270 pounds for the A4.

Consideration of the factors of repetitive manufacture and simplification of maintenance led to the particular arrangements of the A5 bank. Essentially, a channel bank provides twelve distinct circuits. These are basically the same in purpose, but each differs in one respect — operating frequency of the sideband produced. In the A5 design the circuit portions of each channel which are not frequency-sensitive are combined in identical units. Thus there are twelve exactly similar packages for each bank which contain the modulator and demodulator, the voice amplifier, and the level adjusting pads. Fig. 12 shows this "modem" unit. This twelve-fold increase in repetitive manufacture of a major portion of the bank leads to manufacturing simplification and economy. Also, for the first time it will be possible for the Operating Telephone Companies to stock a few spares of the active units which can be used in any channel.

As illustrated by Fig. 13, ease of installation and maintenance was carefully considered in the A5 design. Essentially the bank is comprised of two main sections. The main mounting frame contains the twelve channel filters and the common units such as the hybrids and the network which improves the operation of the end channels. All of these units are removable from the front of the assembly. Simple terminal strips are provided on both sides of this frame for the input and output cables. These are arranged for solderless wire-wrapping.

Attached to this frame is a hinged panel carrying the twelve identical "modem" units. Connection to these is by multi-pin connectors, and the units themselves are easily demountable. Thus, any "modem" unit can be removed from a channel without disturbing other working units. Seen on the rear of the door are small inductors which form part of the terminating filters for the modulators and demodulators. Since these differ for certain groups of channels they, too, have been isolated from

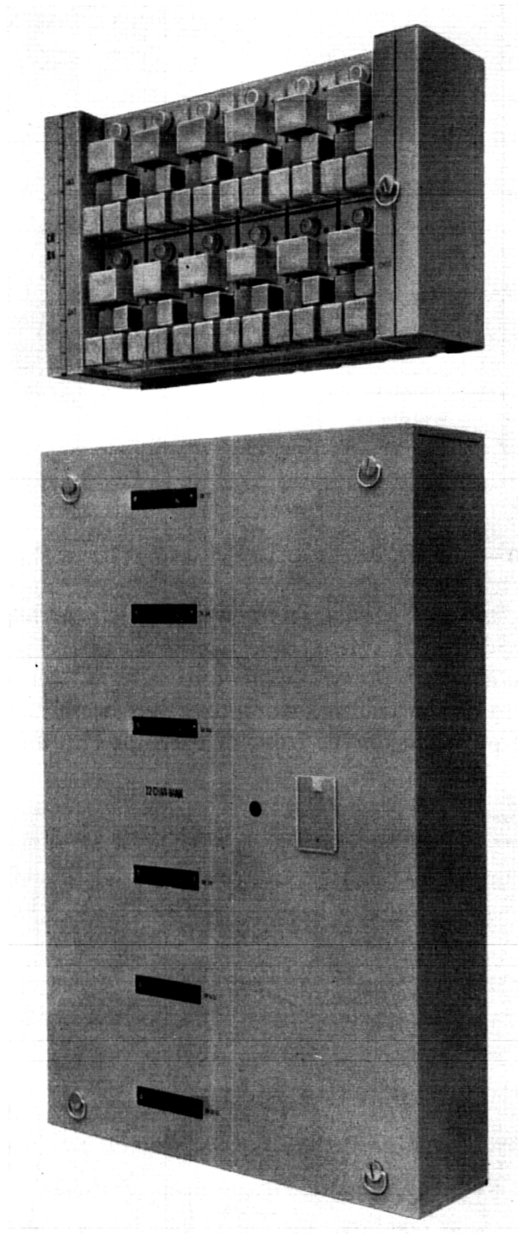

Fig. 9 — Comparison of A5 and A4 channel banks.

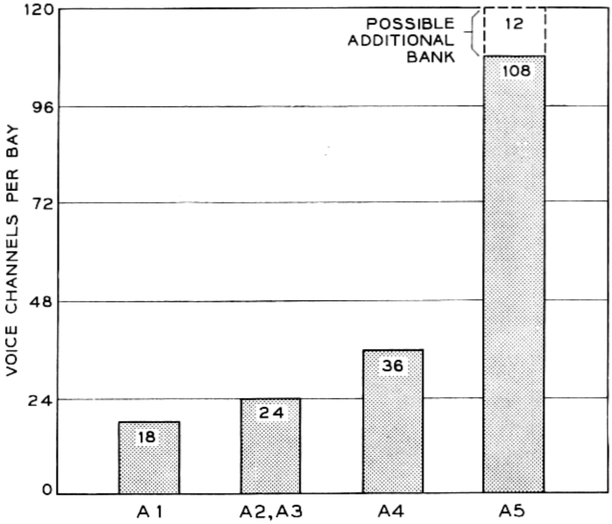

Fig. 10 — Size reductions in A-type channel banks since the A1.

the common "modem" units. The new equipment arrangements of the A5 bank provide for easy front-side maintenance and make even more attractive the common practice of mounting bays back to back.

Maintenance of the transistor amplifier has been made very simple. Pin jacks are provided on the front of each modem unit which permit

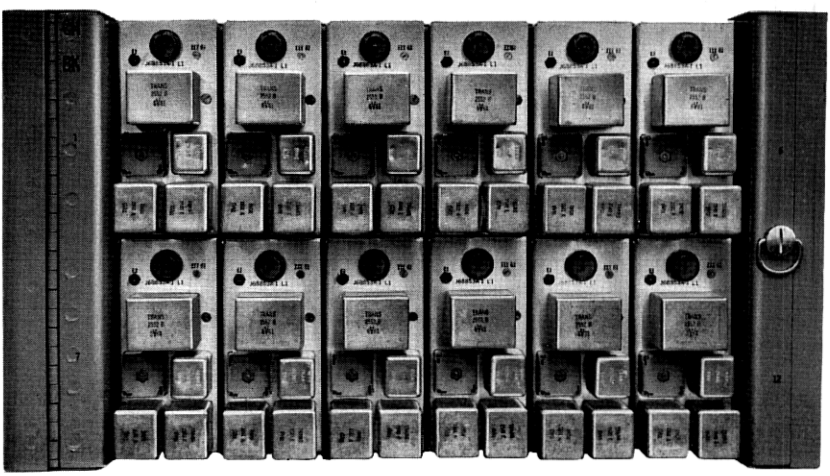

Fig. 11 — A5 bank with hinged panel closed.

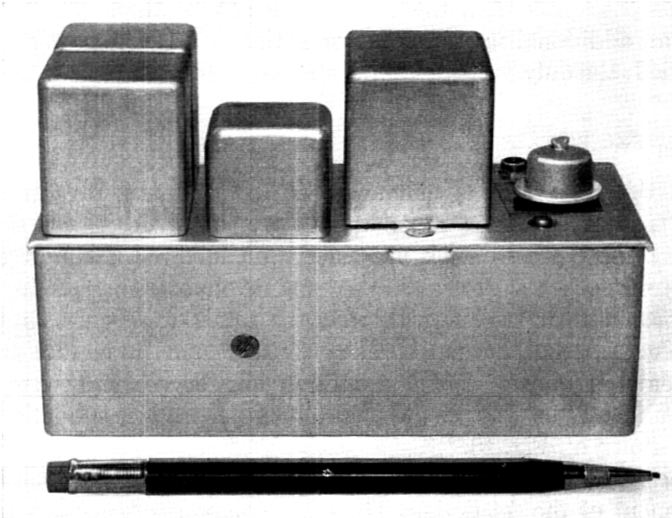

Fig. 12 — Modem unit of A5 bank.

measurements to indicate any degradation in the transistors. With the expected very long life of transistors, it is hoped that field experience will prove that no maintenance tests of this kind are needed. In this event it is likely that the pin jacks would no longer be provided.

From a power basis, the A5 bank offers substantial operational savings. It dissipates only 17 watts as compared to the 59 watts of an A4 bank, better than a 3:1 reduction. This means that even with miniaturization, there is no heat problem. A full complement of nine or ten A5

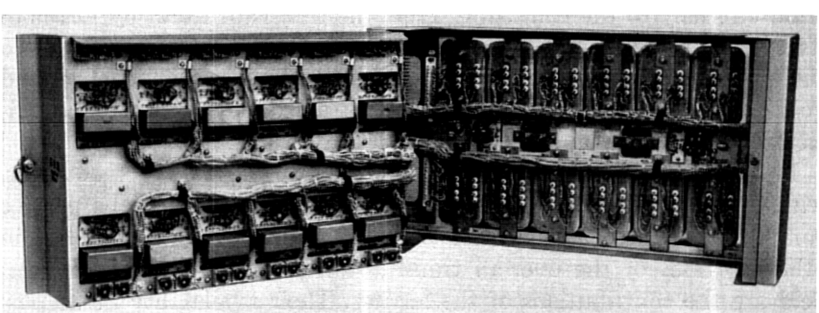

Fig. 13 — A5 bank with hinged door open showing rear of modem units and the filters mounted on rear frame.

units dissipates less than the present full bay of three A4 banks. Of course, an additional favorable factor is that a 130-volt power plant is not needed; the only voltage used is standard office 24 volts.

VII. CHANNEL FILTERS

From the very start of channel bank development, its filters have been a most important component. As previously described, they actually set the pattern which has since been followed. Each of the redesigns from A1 to A4 were also based mainly on filter changes, including reductions in their size. For the A1, the filters were quite large and expensive.[14] Later filters based on newly developed inductors and improvements in quartz crystal mounting and encasement were much smaller and led the way to the over-all equipment size reductions previously shown on Fig. 10.[20]

In the A5 development the filters represent mainly a packaging rearrangement of the A4 design. The main components, crystals and inductors, had already been very substantially reduced in size. To obtain minimum volume, the transmitting and receiving filters, each having the configuration shown in Fig. 14, are combined in one container. By careful placement of the two units and the wiring, and by individual shielding of the inductors, acceptable crosstalk coupling has been achieved. The filter mechanical arrangements are shown on Fig. 15.

Suppression characteristics of a typical filter are shown on Fig. 16 and the performance over the transmission band on Fig. 17.

VIII. OTHER COMPONENTS

The impressive reduction in size of the channel bank cannot be attributed solely to the transistor. True, the transistor itself, by virtue of its over-all size and low power consumption, contributed markedly to the miniaturization of the active circuitry. And the low impedances and voltages of the transistor amplifier circuit aided, as, of course, did efficient equipment design.

Full advantage could not have been taken of these factors, however, without the parallel advances that had been made in various passive components. These are effective not only in the active circuit but in other portions of the over-all transmission path. Outstanding in this field are the contributions of the ferrites. Their availability made possible extreme reductions in the size of transformers in all categories. Fig. 18 shows a comparison between some of the old and new units.

Similar reductions were achieved in the capacitor field. Small Mylar units are used in many instances. In addition, in the amplifier circuitry the very high capacitance values demanded by the low-impedance levels are furnished by miniature solid tantalum units. Without the latter the miniature equipment design would not have been possible.

IX. OVER-ALL PERFORMANCE

The over-all performance of the A5 channel bank meets its development objectives. For convenience in depicting the various characteristics they are discussed under two general headings: (1) those factors which were considered satisfactory in the earlier banks and are essentially unchanged, (2) those in which improvement was sought.

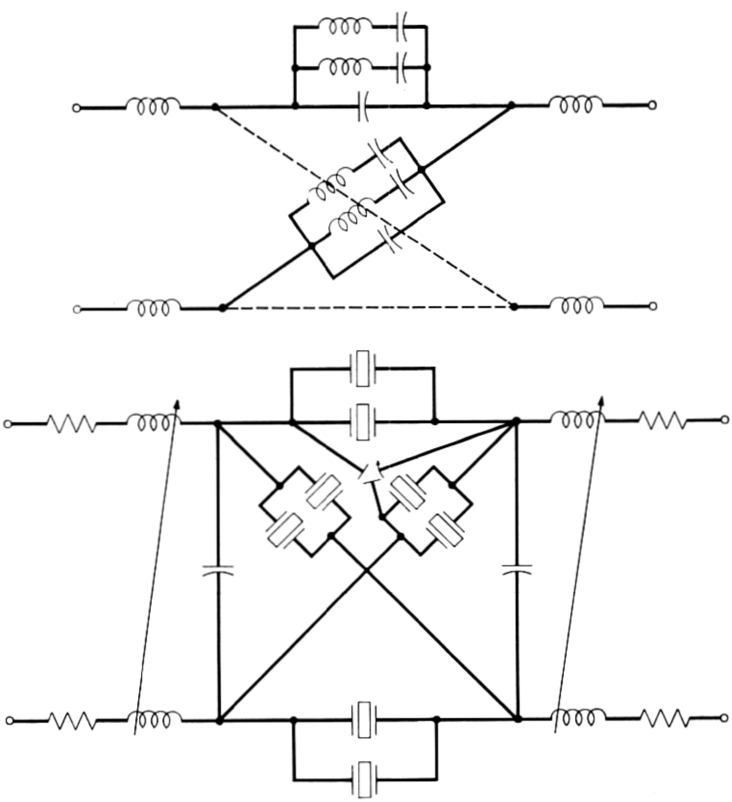

Fig. 14 — Configuration of a crystal channel filter.

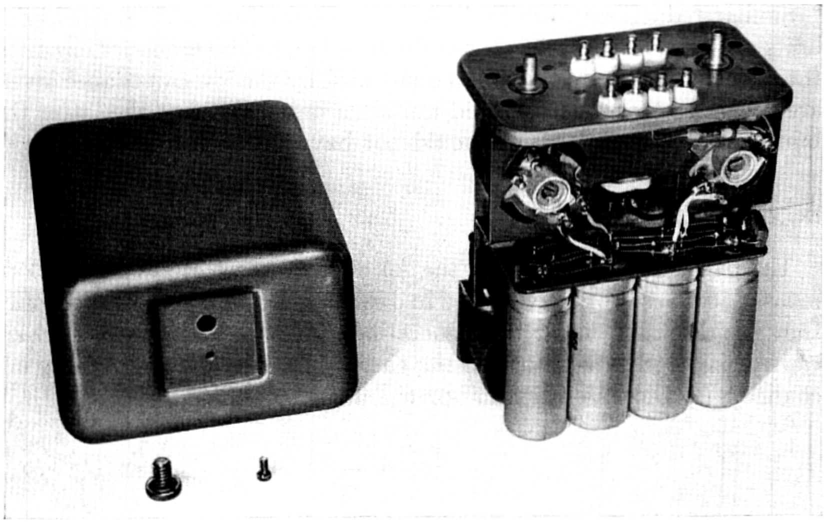

Fig. 15 — Crystal channel filter assembly comprising a transmitting and a receiving unit.

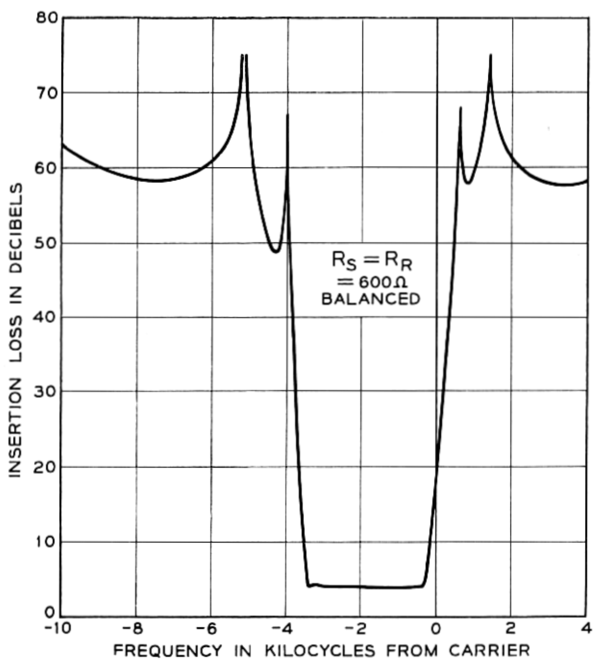

Fig. 16 — Typical loss-frequency characteristic of a channel filter.

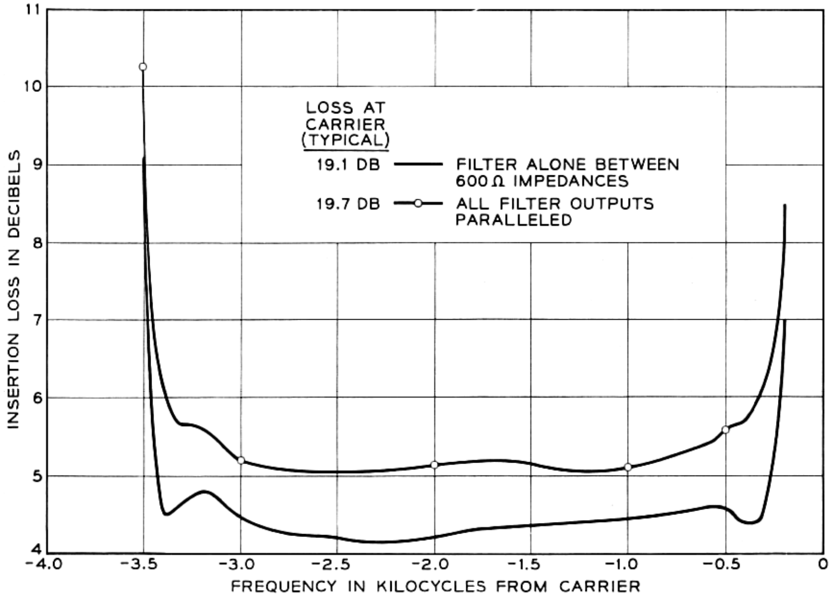

Fig. 17 — Typical pass-band characteristic of a channel filter.

In the first category are frequency response and modulator limiting. Since the chief determinants of the frequency response are the channel band filters, and since they are unchanged in electrical design, the same frequency performance is to be expected. A typical over-all channel characteristic is shown on Fig. 19. The gain-frequency behavior of a demodulator amplifier is shown on Fig. 8.

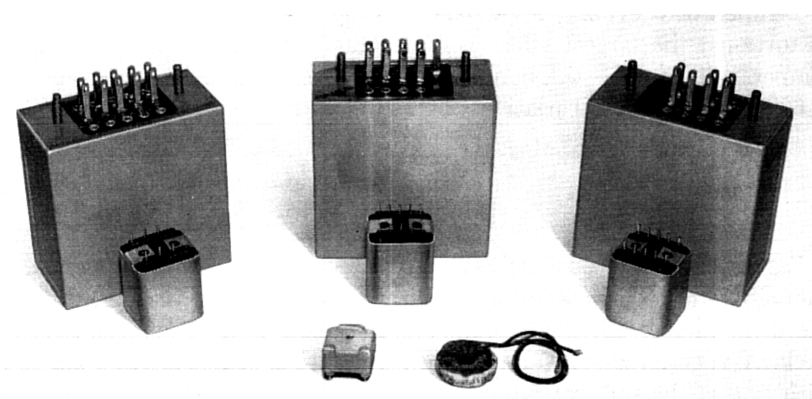

Fig. 18 — Comparison of transformers and inductors used in A5 and A4 banks.

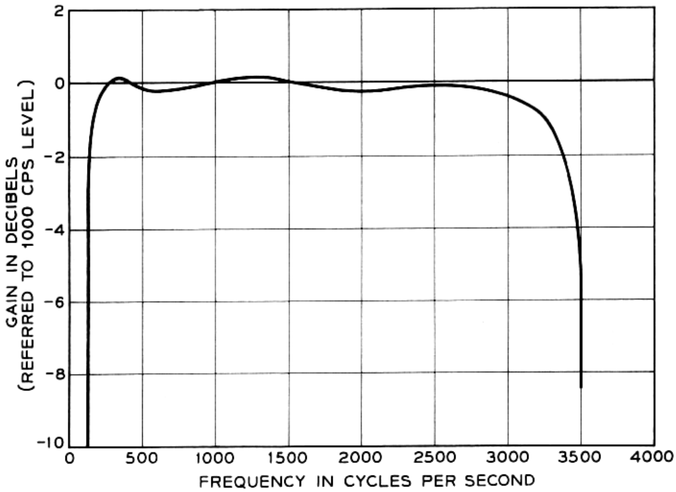

Fig. 19 — Typical channel gain-frequency characteristic of A5 bank.

The channel envelope delay is shown on Fig. 20. In terms of delay distortion, the following is derived from the curve:

Frequency Range (cycles per second)	Delay Distortion (microseconds)
1000–2500	100
850–2700	200
750–2900	300
600–3100	500

Modulator limiting is the same for both A5 and A4 and is shown on Fig. 21.

Temperature cycling tests indicate only very slight effect on band distortions; the largest effect is on channel net loss. From a nominal temperature of 80°F the bank was subjected to variations as great as ±60°F. The net loss variations are as follows:

Temperature Swing	Net Loss Change
±20°F	0.05 db
±40°F	0.25 db
±60°F	0.50 db

Earlier this paper outlined certain desired improvements in operating characteristics. These were particularly concerned with the behavior of the demodulator amplifier, primarily due to its lack of sufficient feedback. A very important aim was the stabilizing of net loss. In the A4 with normal battery variations, changes of two or three db were not uncommon. This situation was, of course, aggravated by aged tubes

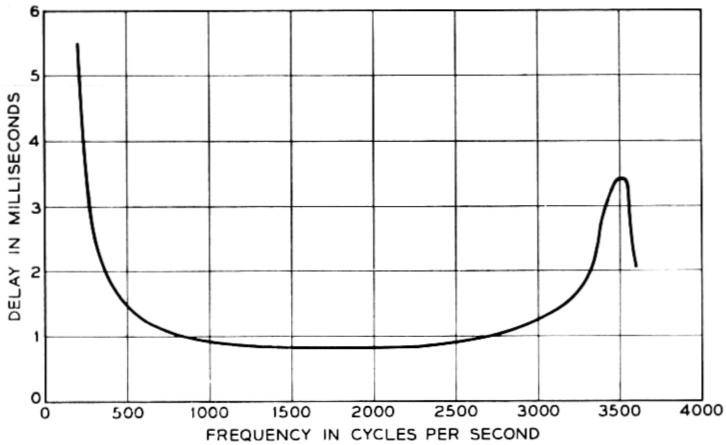

Fig. 20 — Typical channel delay-frequency characteristics of A5 bank.

and maximum gain settings. With 30-db minimum feedback, the aging of active elements should have negligible effect. Also, tests indicate that battery variations cause changes of only about 0.01 db per volt.

A4 channel banks require careful placing with regard to 60-cycle power sources in order to reduce noise pickup. The trouble arises in the amplifier input transformer which has a very high-impedance secondary and a large air gap. In the A5, the transistor circuitry requires a low-impedance transformer with a small air gap. Noise tests indicate that the pickup has been reduced to almost unmeasurable levels.

The intermodulation products created in previous banks make them

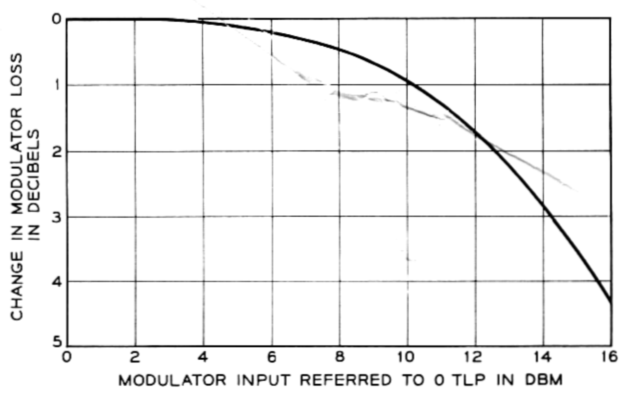

Fig. 21 — Typical channel overload characteristic showing modulator limiting.

unusable for certain special services, such as telephoto, without extreme lowering of levels and the addition of special amplifiers. The A5 amplifier has greatly improved this characteristic. Measurements show that for 0-dbm outputs of each of two frequencies, the $A - B$ product is -63 dbm and the $2A - B$ is -59 dbm. The requirement for telephoto, for example, is -48 dbm.

The amplifier, as designed to meet all of the requirements and employing the 9D power transistor, has a power-handling capacity superior to the older amplifier by about 6 db. The output characteristic as shown on Fig. 22 indicates a break point at about 22 to 23 dbm.

X. CONCLUSION

The A5 channel bank introduces the transistor into the long-haul wire and radio plant of the Bell System. Undoubtedly it is the forerunner of transistor circuitry in other portions of this equipment.

At about the same first cost of equipment, the A5 bank provides definitely improved service to meet the needs of today's communications. In the over-all it means savings to the Operating Telephone Companies through its reduced space and power requirements and its easier maintenance. With the expected high reliability and long life of transistors, the new bank should give many years of service without replacement of its active elements.

XI. ACKNOWLEDGEMENTS

A development of the magnitude of the A5 bank represents the contributions of many people. Certain ones played key roles, and the au-

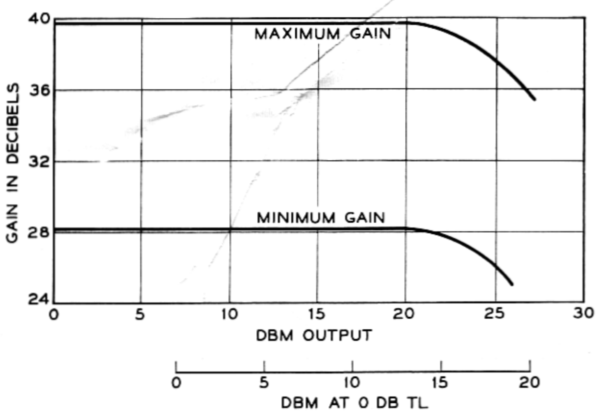

Fig. 22 — Overload characteristic of the transistor amplifier at 1 kc.

thors wish to acknowledge their very substantial part in the development. These are: W. G. Albert, F. R. Bies, J. L. Donoghue, J. B. Evans and J. J. Ginty of Systems Development, Merrimack Valley Bell Laboratories; W. E. Ballentine of Systems Development, Murray Hill Bell Laboratories; and H. G. Wells, now of the Western Electric Co., Merrimack Valley. Also, they wish to thank R. C. Boyd and J. J. Mahoney, Jr. of Systems Engineering for their co-operation and many helpful suggestions. To all the many others, unnamed, the authors also express deep appreciation.

APPENDIX A

Voltage Gain and Feedback Analysis for Transistor Amplifier

A.1 *Series Feedback Amplifier*

In this part of Appendix A, the expressions for voltage gain, feedback and input and output impedances for the series feedback amplifier shown in Fig. 2 are derived. It will be assumed that the amplifier A has no internal feedback and that the Z-parameter matrix for the amplifier is equal to

$$\begin{bmatrix} Z_{\mathrm{IN}}' & 0 \\ Z_{21} & Z_{\mathrm{OUT}}' \end{bmatrix} \tag{12}$$

where Z_{IN}' and Z_{OUT}' are the input and output impedances of the amplifier (without external feedback) respectively. The short-circuit current gain of the amplifier is related to the elements of this matrix by the expression

$$G_i = -\frac{Z_{21}}{Z_{\mathrm{OUT}}'}. \tag{13}$$

The Z-parameter matrix for the feedback network is

$$\begin{bmatrix} Z_{11F} & Z_{12F} \\ Z_{12F} & Z_{22F} \end{bmatrix} \tag{14}$$

where

$$Z_{11F} = \frac{R1 \cdot (R2 + Z2)}{R1 + R2 + Z3}$$

$$Z_{12F} = \frac{R1 \cdot Z3}{R1 + R2 + Z3}$$

$$Z_{22F} = \frac{Z3 \cdot (R1 + R2)}{R1 + R2 + Z3}.$$

Since the amplifier A and the feedback network are connected in series, the over-all matrix for the series feedback amplifier is

$$\begin{bmatrix} Z_{\text{IN}}' + Z_{11F} & Z_{12F} \\ Z_{21} + Z_{12F} & Z_{\text{OUT}}' + Z_{22F} \end{bmatrix}. \tag{15}$$

The voltage gain of the series feedback amplifier is equal to

$$G_v = \frac{(Z_{21} + Z_{12F})R_L}{(Z_{\text{IN}}' + Z_{11F} + R_G)(Z_{\text{OUT}}' + Z_{22F} + R_L) - Z_{12F}(Z_{21} + Z_{12F})}. \tag{16}$$

At this point in the analysis it is convenient to introduce the loop current transmission, which is a convenient measure of feedback for a transistor amplifier. The current transmission is evaluated by opening the feedback circuit at terminals 1-1′, terminating the left-hand pair of terminals in an impedance equal to Z_{IN}' and applying a unit input current to the right-hand pair of terminals. The loop current transmission is equal to the current in Z_{IN}'. The positive direction for this current is chosen so that if the original feedback circuit is restored, the current flows in the same direction as the unit input current. By straightforward calculation

$$A\beta = -\frac{G_i Z_{\text{OUT}}' Z_{12F}}{(Z_{\text{IN}}' + Z_{11F} + R_G)(Z_{\text{OUT}}' + Z_{22F} + R_L) - Z_{12F}^2} \tag{17}$$

$$\frac{A\beta}{1 - A\beta} = -\frac{G_i Z_{\text{OUT}}' Z_{12F}}{(Z_{\text{IN}}' + Z_{11F} + R_G)(Z_{\text{OUT}}' + Z_{22F} + R_L) - Z_{12F}^2 + G_i Z_{\text{OUT}}' Z_{12F}}. \tag{18}$$

If (13) and (18) are substituted into (16), then

$$G_v = \left[\frac{R_L}{Z_{12F}} - \frac{R_L}{G_i Z_{\text{OUT}}'}\right] \cdot \left[\frac{A\beta}{1 - A\beta}\right]. \tag{19}$$

In practice $|G_i| \gg 1$ and $|Z_{\text{OUT}}'| \gg |Z_{12F}|$. To a good approximation

$$G_v = \frac{R_L}{Z_{12F}} \cdot \frac{A\beta}{1 - A\beta} \tag{20}$$

The expressions for input and output impedances for the series feedback amplifier are readily obtained using Blackman's formula.[23]

$$Z = \frac{Z_0(1 - A\beta_0)}{(1 - A\beta_\infty)}. \tag{21}$$

In this formula, Z is the driving point impedance between two points in a circuit, Z_0 is the driving point impedance when one of the transistors in the amplifier A is in a reference condition so that $G_i = 0$ (feedback circuit is opened), $A\beta_0$ is the loop current transmission when the two points between which the impedance is measured are shorted, and $A\beta_\infty$ is the loop current transmission when the two points are open circuited.

In the case of input impedance, $Z_0 = (Z_{\text{IN}}' + Z_{11F})$, the input impedance of the amplifier A with the feedback loop opened plus Z_{11F}. $A\beta_0$ is equal to the loop current transmission with the input terminals to the series feedback amplifier, 1-3′, shorted. This value of loop current transmission can be calculated from (17) with R_G set equal to zero, and will be designated as $A\beta(R_G = 0)$. $A\beta_\infty$ is equal to the loop current transmission with the input terminals, 1-3′, open. Obviously $A\beta_\infty = 0$.

$$Z_{\text{IN}} = (Z_{\text{IN}}' + Z_{11F})[1 - A\beta(R_G = 0)]. \tag{22}$$

Similarly the expression for output impedance measured between terminals 2-4′ is

$$Z_{\text{OUT}} = (Z_{\text{OUT}}' + Z_{22F})[1 - A\beta(R_L = 0)] \tag{23}$$

where Z_{OUT}' is the output impedance of the amplifier A with the feedback loop opened and $A\beta(R_L = 0)$ is equal to (17) with R_L set equal to zero.

A.2 *Emitter Feedback Amplifier*

The emitter feedback amplifier shown in Fig. 3 differs from the series feedback amplifier in that the emitter of the second transistor and the interstage networks are returned to the common connection between the input and output transformers instead of to the emitters of the first and third transistors. As a result of this connection, the collector current of the first transistor must pass through the feedback network ($R1$, $R2$ and $Z3$) in returning to the emitter and, therefore, the network introduces local feedback to the first transistor stage. This feedback is present even when the main feedback loop is opened. Similarly, the base current of the third transistor stage must pass through the feedback network in returning to the emitter. The feedback network, consequently, introduces feedback to the third transistor stage.

The emitter feedback amplifier also differs from the series feedback amplifier in that the feedback network is driven by the emitter current of the output stage instead of by the collector current. Since the small signal collector current of a junction transistor is equal to the small signal emitter current multiplied by the alpha of the transistor, expression (20) must be modified for the emitter feedback amplifier.

$$G_v = \frac{R_L a_3}{Z_{12F}} \cdot \frac{A\beta}{1 - A\beta}. \tag{24}$$

In order to calculate the feedback developed by an emitter feedback amplifier, it is convenient to use the circuit shown in Fig. 4. This circuit, to a good approximation, takes into account the local feedback introduced by the feedback network. Expressions (17), (22) and (23) are valid for the emitter feedback amplifier if Z_{IN}' and Z_{OUT}' are calculated for the circuit shown in Fig. 4.

APPENDIX B

Low-Frequency Stability Analysis of Transistor Amplifier

All of the low-frequency feedback shaping is introduced by the three capacitors in the emitter circuits of the transistors (refer to Fig. 5). Fig. 6 shows a plot of the low-frequency loop current transmission. The gain-cutoff at the frequency f_{A3} is introduced by the capacitor C_{E3}, in the emitter circuit of the last stage. At the cutoff frequency, the input impedance of the third stage is equal to the total shunt resistance, R_{S3}, between the base of the transistor and ground. To a good approximation

$$f_{A3} = \frac{1}{2\pi C_{E3} R_{S3}(1 - a_{03} + \delta)} \qquad (25)*$$

where

$$\delta = \frac{R_L}{r_{c3}}.$$

The cutoff introduced by C_{E3} is terminated at the frequency f_{B3} at which the reactance of C_{E3} is equal to R_{E3}.

$$f_{B3} = \frac{1}{2\pi C_{E3} R_{E3}}. \tag{26}$$

The gain-cutoff at the frequency f_{A2} is introduced by C_{E2}. At this cutoff frequency, the input impedance of the second stage is equal to

* The first subscript for corner frequencies refers to the type of cutoff, while the second subscript refers to the transistor stage.

the total shunt resistance, R_{S2}, between the base of the transistor and ground. To a good approximation

$$f_{A2} = \frac{1}{2\pi C_{E2} R_{S2}(1 - a_{02})}. \tag{27}$$

The cutoff introduced by C_{E2} is terminated at the frequency f_{B2} at which the reactance of C_{E2} is equal to R_{E2}.

$$f_{B2} = \frac{1}{2\pi R_{E2} C_{E2}}. \tag{28}$$

With reference to (9), it is evident that C_{E1} will introduce a gain cutoff at the frequency where the reactance of the capacitor is equal to

$$[(r_{b1}' + Z_{11F} + R_G)(1 - a_{01}) + r_{e1} + Z_{11F}]$$

$$f_{A1} = \frac{1}{2\pi C_{E1}[r_{b1}' + R_{11F} + R_G)(1 - a_{01}) + r_{e1} + R_{11F}]}. \tag{29*}$$

This cutoff is terminated at the frequency f_{B1} at which the reactance of C_{E1} is equal to R_{E1}.

$$f_{B1} = \frac{1}{2\pi R_{E1} C_{E1}}. \tag{30}$$

APPENDIX C

High-Frequency Stability Analysis of Transistor Amplifier

In this Appendix we will calculate the high-frequency characteristic of the loop current transmission. Expression (17) is valid for the emitter feedback amplifier if Z_{IN}' and Z_{OUT}' are calculated for the circuit shown in Fig. 4.

$$G_i = \frac{a_1}{1 - a_1} \cdot \frac{a_2}{1 - a_2} \cdot \frac{1}{1 - a_3} \tag{31}$$

$$Z_{\text{IN}}' = r_{b1}' + \frac{r_{e1} + Z_{11F}}{1 - a_1} \tag{32}$$

$$Z_{\text{OUT}}' = Z_{c3}(1 - a_3) + Z_{22F}. \tag{33}$$

In practice, the term Z_{22F} in the expression for Z_{OUT}' can be neglected.

* At the frequency f_{A1}, Z_{11F} is real and has the value R_{11F}.

If (31), (32) and (33) are substituted into (17), then

$$A\beta = -\frac{\dfrac{a_1 a_2 Z_{12F}}{1 - a_2}}{[r_{e1} + Z_{11F} + (r_b' + Z_{11F} + R_G)(1 - a_1)] \cdot \left[(1 - a_3) + \dfrac{R_L + Z_{22F}}{Z_{c3}}\right] - \dfrac{Z_{12F}^2(1 - a_1)}{Z_{c3}}}. \tag{34}$$

In all practical designs, the term $Z_{12F}^2(1 - a_1)/Z_{c3}$ in the denominator of (34) can be neglected. In the following high-frequency analysis, it is assumed that the transistor parameters a and Z_c have the following frequency characteristic

$$a = \frac{a_0 \exp\left(-j\dfrac{fm}{f_a}\right)}{1 + j\dfrac{f}{f_a}} \qquad (35)^{24}$$

$$Z_{c3} = \frac{r_{c3}}{1 + j2\pi f r_{c3} C_{c3}} \tag{36}$$

where a_0 is the dc value of the device parameter a, f_a is the frequency at which the magnitude of a is 3 db below its dc value, m is the number of radians by which the phase shift of a exceeds $\pi/4$ (45°) at f_a, C_{c3} is the collector capacitance and r_{c3} is the collector resistance of the third transistor. In practice, m is of the order of 0.2 for alloy types of transistors. If expressions (35) and (36) are substituted in (34), then to a good approximation

$$A\beta = \frac{A\beta_0\left(1 + j\dfrac{f}{f_{a3}}\right)}{\left(1 + j\dfrac{f}{f_{11}}\right)\left(1 + j\dfrac{f}{f_{12}}\right)\left(1 + j\dfrac{f}{f_{13}}\right)} \qquad (37)^*$$

where $A\beta_0$ is equal to the mid-band value of $A\beta$ given by (9),

$$f_{11} = \frac{f_{a1}\left[1 - a_{01} + \dfrac{r_{e1} + R_{11F}}{r_{b1}' + R_{11F} + R_G}\right]}{1 + a_{01}m_1 + \dfrac{r_{e1} + R_{11F}}{r_{b1}' + R_{11F} + R_G}}$$

$$f_{12} = f_{a2}(1 - a_{02})$$

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* At high frequencies, Z_{11F} , Z_{12F} and Z_{22F} are real with values R_{11F} , R_{12F} and R_{22F} respectively.

$$f_{13} = \frac{(1 - a_{03} + \delta)}{\dfrac{1 + a_{03}m_3 + \delta}{f_{a3}} + \dfrac{1}{f_{c3}}}$$

$$f_{c3} = \frac{1}{2\pi(R_L + R_{22F})C_{c3}}, \qquad \delta = \frac{R_L + R_{22F}}{r_{c3}}.$$

Expression (37) represents the high-frequency behavior of the loop current gain of the circuit without high-frequency shaping. In order to insure adequate margins against instability, two interstage high-frequency shaping networks are employed as shown in Fig. 5. These networks modify the high-frequency current gain of the second and third stages. In Ref. 21 it is shown that if a series RC circuit is placed in the base circuit of a transistor ($R2$, $C2$) the current gain of the stage can be represented by the following expression:

$$G_i = \frac{\dfrac{a_{02}}{1 - a_{02}}\left(1 + j\dfrac{f}{f_{32}}\right)}{\left(1 + j\dfrac{f}{f_{22}}\right)\left(1 + j\dfrac{f}{f_{42}}\right)} \tag{38}$$

where

$$f_{22} = \frac{1}{2\pi\left[r_{b2}' + R2 + \dfrac{r_{e2}}{1 - a_{02}}\right]C2}$$

$$f_{32} = \frac{1}{2\pi R2\cdot C2}$$

$$f_{42} = \frac{f_{12}\left[r_{b2}' + R2 + \dfrac{r_{e2}}{1 - a_{02}}\right]}{r_{b2}' + R2}.$$

The corner frequency f_{22} corresponds to the frequency at which the reactance of $C2$ is equal to the input impedance of the second transistor stage. The corner frequency f_{32} is equal to the frequency at which the reactance of $C2$ is equal to $R2$.

In Ref. 21 it is shown that if a series RLC circuit is placed in the base circuit of a common emitter transistor ($R3$, $L3$ and $C3$), the current gain of the stage can be represented by the following expression:

$$G_i = \frac{\dfrac{a_{03}}{1 - a_{03} + \delta}\left(1 + j\dfrac{f}{f_{53}}\right)^2}{\left(1 + j\dfrac{f}{f_{23}}\right)\left(1 + j\dfrac{f}{f_{43}}\right)\left(1 + j\dfrac{f}{f_{63}}\right)} \tag{39}$$

where

$$f_{23} = \frac{1}{2\pi\left[r_{b3}' + R3 + \dfrac{(r_{e3} + R_{22F})(1 + \delta)}{1 - a_{03} + \delta}\right]C3}$$

$$f_{43} = \frac{f_{13}\left[r_{b3}' + R3 + \dfrac{(r_{e3} + R_{22})(1 + \delta)}{1 - a_{03} + \delta}\right]}{r_{b3}' + R3}$$

$$f_{53} = \frac{1}{2\pi\sqrt{L3C3}}$$

$$f_{63} = \frac{r_{b3}' + R3}{2\pi L3}.$$

The corner frequency f_{23} corresponds to the frequency at which the reactance of $C3$ is equal to the input impedance of the third transistor stage. The corner frequency f_{43} corresponds to the frequency at which the magnitude of the input impedance of the third stage plus $R3$ is within 3 db of $(r_{b3}' + R3)$. The corner frequency f_{53} corresponds to series resonance of the RLC circuit and f_{63} is the frequency at which the reactance of $L3$ is equal to $(r_{b3}' + R3)$. In order for (39) to be valid, $R3$ in the series resonant circuit must be chosen so that the circuit has a Q of one-half at f_{53}.

$$R3 = \frac{1}{\pi f_{53} C3}. \tag{40}$$

The complete expression for the high-frequency characteristic of $A\beta$ for the amplifier shown in Fig. 5, is

$$A\beta = \frac{A\beta_0\left(1 + j\dfrac{f}{f_{32}}\right)\left(1 + j\dfrac{f}{f_{53}}\right)^2\left(1 + j\dfrac{f}{f_{a3}}\right)}{\left(1 + j\dfrac{f}{f_{11}}\right)\left(1 + j\dfrac{f}{f_{22}}\right)\left(1 + j\dfrac{f}{f_{42}}\right)\cdot\left(1 + j\dfrac{f}{f_{23}}\right)\left(1 + j\dfrac{f}{f_{43}}\right)\left(1 + j\dfrac{f}{f_{63}}\right)}. \tag{41}$$

The magnitude and phase of $A\beta$ is plotted in Fig. 7 for the transistor parameter values listed in Table I and for $|Z_{12F}| = 7$ ohms.

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