A Multi-Loop, Self-Balancing Power Amplifier*

J. R. MACDONALD†

Summary—A multi-loop, push-pull power amplifier of exceptional characteristics is described. It employs special circuits to maintain accurate push-pull signal balance throughout and to hold the static or steady-state dc cathode currents of the output tubes equal. A pair of 807 tubes are used in class AB_2 to yield 65w average power output at less than 1 per cent intermodulation distortion with 30 db of over-all negative feedback. Using local positive voltage feedback in addition, the intermodulation distortion is 0.1 per cent at 45w and less than 0.2 per cent at 60w. At full power output, the −0.5 db points occur at 19.8 cps and 22.4 kcps. The rise time of the amplifier is 3 μs, and its transient response and recovery from overload approach the ideal. There are no peaks at the ends of the response curve. A noise level of −106 db referred to 60w output is attained. Extensive measurements of amplifier characteristics under various conditions are described.

INTRODUCTION

A POWER AMPLIFIER is generally required to supply maximum power output at minimum distortion over a specified bandwidth. In addition, cost and complexity must be relatively low for most commercial applications, and power efficiency should be as high as practical. The design and construction of the present amplifier was begun four years ago. By designing the amplifier without stringent cost and complexity restrictions, it was felt that out-of-the-ordinary characteristics might be attained. Some of the means of achieving such characteristics might then be directly applicable to simpler, lower-cost amplifiers; all of them, it was felt, might yield a useful perspective on what kind of a system could be achieved for a given level of cost and complexity.

The amplifier has been operating in substantially its present form for two years. Since it is a developmental unit, it is not really completed, however. Its complexity and corresponding flexibility are such that it is expected that its present form will not remain completely static.

The initial design goals of the amplifier were as follows:

1. Push-pull operation with a pair of automatically balanced output triodes operating with fixed or automatic bias.
2. A gain of the order of 30 db, and a frequency response variable by no more than one db between 10 and 30,000 cps.
3. A distortion not exceeding 1 per cent intermodulation at maximum power output.
4. Excellent transient response at all levels.
5. "Undistorted" maximum power output over substantially the entire working frequency range.
6. Negligible noise and hum.

All the above goals have been met or exceeded by the amplifier in its present state.

* Manuscript received, February 21, 1955. Paper delivered at Dallas IRE Section, October 21, 1954; Houston IRE Section, March 22, 1955.

† Texas Instruments, Inc., Dallas, Tex.

SIMPLIFIED BLOCK DIAGRAM

Since the complete amplifier is rather complex, it will prove convenient to analyze it section by section. The simplified block diagram of Fig. 1 shows that there are three gain stages. The first stage is a special phase inverter with automatic dynamic balancing. This stage is followed by cathode followers included to extend the frequency response and to serve as low-impedance sources for feedback voltage. Next come the only capacitors in the direct signal path. The gain stage A_2 also incorporates dynamic self-balancing. It is direct-coupled to special drivers of about two ohms output impedance which are directly connected to the output tube grids. Over-all feedback which is an adjustable combination of negative voltage and positive or negative current signals is taken from the secondary of the output transformer and returned to the input. The feedback circuit is shown symbolically in Fig. 1. Since 30 db or more of over-all negative voltage feedback is used from output to input, the frequency response requirements of the various stages and of the output transformer are rather stringent.

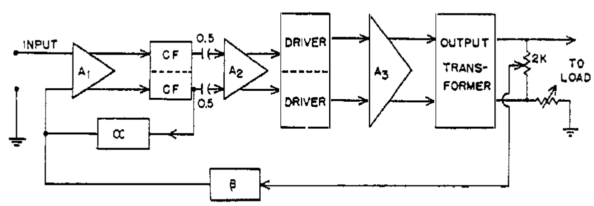

Fig. 1—Simplified block diagram of the amplifier.

PHASE-INVERTER AMPLIFIER STAGE

The initial design of the amplifier employed a cross-coupled phase inverter[1] with the over-all negative feedback returned to one input grid and the other used for the input signal. Although this circuit was found to be satisfactory from inversion and frequency response standpoints, its in-phase response was too high for good operation with a large amount of feedback. When feedback and input signals are applied separately to the two grids of such a circuit, it is desirable that their amplified difference appear at one output plate and the same signal shifted by 180 degrees appear at the other plate. Such behavior was not found.

When feedback signals representing 20 or 30 db of over-all feedback are employed, the difference between the input and the feedback signals is 10 per cent or less, a small difference compared to the magnitude of the individual signals. If the circuit does not have good in-phase (common-mode) rejection, the two large, practically equal, input signals may produce in-phase output signals as large or larger than their out-of-phase

[1] J. N. van Scoyoc, "A cross-coupled input and phase-inverter circuit," *Radio News*, vol. 40, p. 6; November, 1948.

amplified differences. The output will then consist of only two large, almost equal in-phase signals. The difference between them will represent the desired signal. This matter of in-phase response of push-pull amplifiers has been discussed in some detail by Offner.[2] The effectiveness of the over-all feedback is, of course, entirely dependent upon the precision with which this difference between input and feedback signals is produced, and the presence of a common-mode signal in the phase inverter output definitely degrades performance.

In a push-pull amplifier with appreciable common-mode signal at the outputs of the phase inverter, the in-phase signal may be amplified by the rest of the amplifier and at the worst can cause strong overloading of succeeding stages. Such signals, of course, will not pass through to the secondary of the output transformer because of its in-phase discrimination.[3] Even if the in-phase signals do not cause overloading, however, their presence anywhere in the amplifier can produce increased intermodulation. Therefore, it is of considerable importance to ensure that the outputs of any phase-inverter used with appreciable feedback contain only a negligible amount of common-mode signal.

The unconventional self-balancing phase inverter finally used in the amplifier is shown in Fig. 2. The basic idea for a simplified version of this circuit was suggested to the author by L. C. Labarthe.[4] Since then, it has been found that the same idea has been developed independently.[5] The function of the circuit is as follows. A signal at either input grid produces an amplified output across the cathode of the corresponding cathode follower. The two 100K resistors between the cathode follower outputs are closely matched. If the cathode follower outputs are accurately 180 degrees out-of-phase, no signal will appear at the junction of the resistors. Should any in-phase components be present at the cathode follower outputs, however, an error signal will appear at this junction. This error signal then passes through the cathode-follower voltage divider V_a to the constant-resistance tube V_b.[6] The error signal then finally appears at the plate of V_b where it drives the cathodes of the input double triode in such phase that the error signal at the outputs is itself greatly reduced. It will be seen that the circuit itself therefore uses a kind of negative feedback which has been called "active-error feedback."[7]

Experimentally, it is found that the out-of-phase to in-phase amplification ratio of this circuit is between 10^3 and 10^4. At all frequencies from zero up to more than 500kc, the output signals remain closely 180 degrees out-of-phase, and even with 30 or 40 db feedback no common mode output signal is detectable. Such behavior is largely independent of tube ageing effects since it is produced by feedback. In addition, another important advantage of the circuit is that all even harmonics act like in-phase signals and are themselves automatically greatly reduced in amplitude at output.

The cathode-follower outputs of the phase-inverter-amplifier drive two feedback loops and the succeeding stage. We shall defer until later discussion of all such feedback loops to the input. The 0.5 μf isolating capacitors have about 400v across them. They are, therefore, low-leakage oil-filled units. Their capacity to ground is minimized by mounting them on porcelain stand-off insulators inside the amplifier chassis. The shielding shown in Fig. 2 requires some comment. It will be seen that the shields are also driven by the cathode followers. Such driven shields can do much to extend the very-high-frequency response of the input amplifier stage by reducing its output capacity to ground. The capacity from each plate circuit to its shield is of no importance because the shield is driven by a voltage practically identical to that at the plate.[8] The decoupling networks for the -380v and $+400$v supplies shown in Fig. 2 are not necessary to eliminate motor boating; it was found that their addition markedly decreased the intermodulation distortion of this stage.

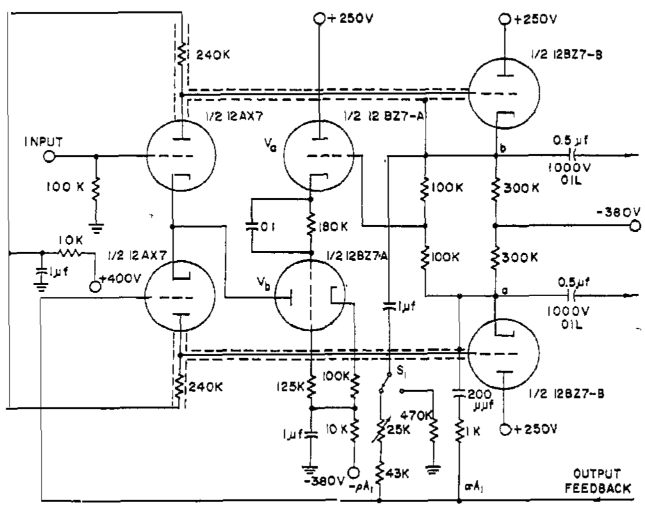

Fig. 2—Phase-inverter amplifier stage circuit.

Driver Amplifier Stage

The driver amplifier stage of Fig. 3 (next page) is quite similar to the input phase-inverter-amplifier stage except it operates with a push-pull input. A dynamic balancing circuit is here employed to accomplish three objectives: stabilization of the average value of the bias voltages applied to the grids of the output tubes; production of stable, accurately push-pull driver signal voltages; and elimination of even-order harmonic distortion in the driver outputs.

[2] F. F. Offner, "Balanced amplifiers," Proc. IRE, vol. 35, pp. 306–310; March, 1947.
[3] The in-phase discrimination may be poor at high frequencies because of capacitative coupling between primary and secondary unless electrostatic shielding is employed.
[4] Private communication in 1951.
[5] E. M. I. Laboratories, "Balanced output amplifiers of highly stable and accurate balance," *Electronic Engrg.*, vol. 18, p. 189; June, 1946.
[6] G. E. Valley and H. Wallman, "Vacuum Tube Amplifiers," McGraw-Hill Book Co., Inc., New York, N. Y., p. 432; 1948.
[7] J. R. Macdonald, "Active-error feedback and its application to a specific driver circuit," Proc. IRE, vol. 43, pp. 808–813; July, 1955.
[8] J. R. Macdonald, "An ac cathode-follower circuit of very high input impedance," *Rev. Sci. Instr.*, vol. 25, pp. 144–147; February, 1954.

The present amplifier incorporates a direct-coupled static balance circuit which acts to keep the average values of the cathode currents of the two output tubes equal. This function, which will be discussed in detail later, is accomplished by means of a closed feedback loop which holds the grid bias voltages of the output tubes to the correct values. The dc and ultra-low-frequency bias correction signals are injected into the grids of the driver amplifiers through frequency-sensitive networks at the points marked *SB* in Fig. 3. These signals are themselves accurately push-pull; therefore, after amplification in the driver amplifier stage, they have no effect on the average value of the output tube grid bias voltages.

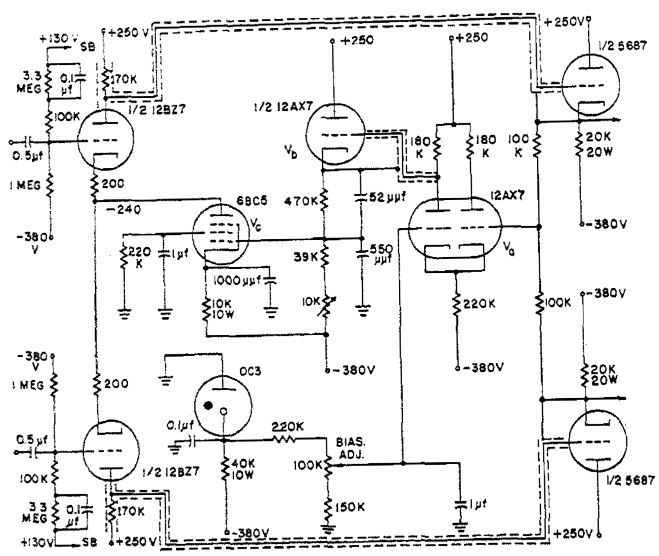

Fig. 3—Driver-amplifier stage circuit.

Part of the modified, direct-coupled, cathode-follower driver circuits are shown in Fig. 3. These special circuits will be discussed in the next section. Their outputs, at the cathodes of the 5687 tube of Fig. 3, go directly to the output tube grids and also, through the matched 100K resistors, to the dynamic balance circuit. The signal at the junction of these resistors is proportional to the average value of the two bias voltages and contains any signal arising from deviations of the driver signals from exact push-pull conditions and any even-order harmonic components of these signals. The average value of the grid bias voltages is stabilized by means of the differential amplifier tube V_a. The voltage on one grid is the desired average bias, adjusted and stabilized by means of the 100K potentiometer and the OC3 regulator tube. The amplified difference between this dc voltage and the actual average bias on the other grid of V_a appears at its plate. This signal, together with other amplified error components, then goes through the frequency-compensated, cathode-follower voltage divider V_b and thence to the constant-current amplifier pentode V_c. The 1,000 $\mu\mu$f cathode bypass capacitor of V_c acts to increase the stage gain at high frequencies. The amplified error signal then passes through the driver amplifier tube halves and the drivers where it re-enters the feedback loop in such phase that it is greatly reduced in magnitude. This direct-coupled feedback loop comprises three gain stages and has a very high gain which is not greatly reduced until the megacycle frequency range is entered. It is, therefore, very effective in eliminating in-phase error components in the driver signals and keeping the average value of the two bias voltages fixed. The shielding shown functions like that of Fig. 2 to improve the high-frequency response of the main amplifier and of the dynamic balance loop. Since the driver amplifier tubes and the series pentode V_c of Fig. 3 operate with their cathodes appreciably below ground potential, they are supplied from a separate, negatively biased heater supply to avoid exceeding their heater-cathode voltage ratings.

The bias adjustment allows the average bias to be adjusted from about -34v to -58v. When reasonably well-matched output tubes are employed, and the static balance circuit used to hold their steady-state cathode currents equal, it is found that their bias voltages differ by less than 0.5v at most over the entire bias and signal voltage ranges.

In an earlier version of the amplifier, the dynamic balance circuit of Fig. 3 was differently arranged so that a signal equal to the sum of a constant voltage and a voltage proportional to the average value of the cathode currents of the output tubes could be used to determine the average value of the output-tube bias voltages. This is a form of automatic bias control so arranged that the magnitude of the average bias is increased as the signal, and so the output cathode currents, increase. This modification was eliminated later after it was found unnecessary from an output distortion viewpoint. Such elimination is desirable, when possible, since the transient reponse of such a circuit is not good. Such automatic bias control should operate as rapidly as possible for good transient response; yet, by definition, its response time must be appreciably slower than the lowest signal frequency of the amplifier. These conditions are somewhat conflicting.

In concluding this section, the desirability of making the signals to the output tube grids accurately push-pull should be emphasized. Accurate push-pull balance at this point can reduce nonlinear distortion in the output tubes appreciably. A number of recent amplifiers[9–11] use balanced push-pull feedback taken from the output tube plates to improve push-pull balance. As Good[12] has pointed out, since the output transformer

[9] B. B. Drisko and R. D. Darrell, "40-db feedback audio amplifier," *Electronics*, vol. 25, pp. 130–132; March, 1952.

[10] J. M. Diamond, "Multiple-feedback audio amplifier," *Electronics*, vol. 26, pp. 148–149; November, 1953.

[11] J. Z. Knapp, "The linear standard amplifier," *Radio and TV News*, vol. 51, pp. 43–46, 113–114; May, 1954.

[12] E. F. Good (Letter), *Electronics*, vol. 25, pp. 420–422; October 1952.

functions as an auto-transformer, the plate voltages will be essentially balanced whether the grid voltages are or not. Thus, such plate feedback is necessarily ineffective in producing output-tube current balance and in reducing unbalance distortion. Although a feedback loop might be arranged to keep the instantaneous output tube plate currents accurately push-pull, the present dynamic balance circuit acting on the grids accomplishes practically the same result (provided the output-tube transconductances are fairly well-matched) and does not reduce amplifier gain as does the ineffective balanced plate feedback.

Augmented Cathode-Follower Driver

In order to obtain maximum power output, it is desirable to drive the grids of the output stage into the grid current region. Such a procedure can increase the output power greatly, but it is eventually limited when the grids are driven positive to the diode line, where they lose control of the plate current. To drive 807 grids to the diode line, peak positive grid currents of the order of 100 to 200 milliamperes are required from the preceding driver stage. In addition, since the input impedance of an output tube grid is a nonlinear function of the grid voltage in the positive grid region, the above large driving currents must be supplied from a source of very low internal impedance to avoid grid-voltage distortion during this part of each cycle.

The output impedance of the driver amplifier of Fig. 3 is quite high; therefore, an impedance converter is required between this stage and the output stage. Since an ordinary direct-coupled cathode-follower driver still has too high an output impedance to be ideal for

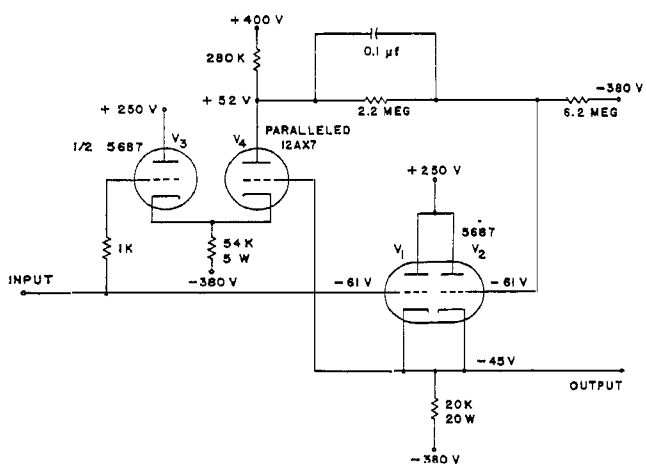

Fig. 4—Direct-coupled driver circuit, one push-pull side only.

such a purpose, the special circuit shown in Fig. 4 has been developed as a virtually distortionless driver for this application. This circuit has been discussed in detail and its operating characteristics compared with other kinds of drivers elsewhere.[7] Therefore, here it need only be mentioned that, like the dynamic balance circuit of Fig. 1, it makes use of active error feedback. The difference between the input to cathode-follower V_1 and its output is amplified in V_4, then injected back into V_2 in such phase as to reduce this difference or error. The circuit is direct-coupled, will operate up into the megacycle range, and has an output impedance of about five ohms. Two such circuits are used in the present amplifier, one for each push-pull side. Since the dynamic balance feedback of Fig. 3 is taken from the outputs of these augmented cathode-follower drivers, this further feedback serves to reduce the actual small-signal output impedances of the drivers to between 2 and 3 ohms. For large positive output currents, the impedance is even further reduced to about one ohm by the increase of the g_m's of the output 5687 tubes. The driver can supply peak positive currents of several hundred milliamperes at 50 to 100v rms per side with little or no distortion.[7]

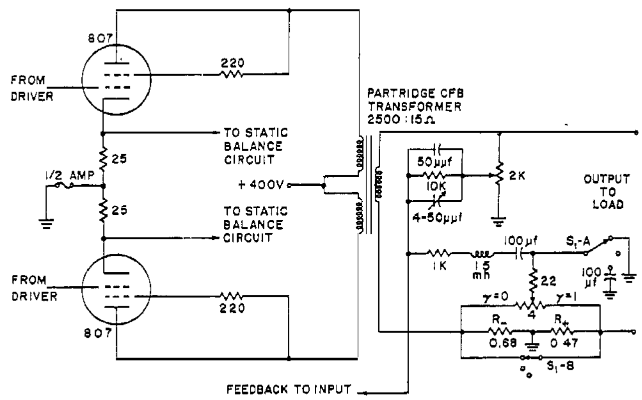

Fig. 5—Output stage circuit including over-all feedback arrangements.

Output Stage

The output stage shown in Fig. 5 uses triode-connected 807 tubes and a high-quality, commercial grain-oriented output transformer rated at 60w between 30 and 30,000 cps with less than 1 per cent total harmonic distortion without feedback. This transformer has eight secondary sections and offers a wide selection of output impedances, but only the 15-ohm output is used in the present design. It has very low leakage inductance between the primary half-sections and hence is suitable for class B operation. Since the variable bias control of Fig. 3 allows the bias on the 807 grids to be adjusted over a wide range, the class of operation of the output tubes may lie anywhere from A_2 to B_2.

For longest tube life and minimum distortion, it is desirable that the quiescent plate currents of the output tubes be equal. Unequal currents tend to saturate the output transformer, with consequent degradation of low-frequency response. The 25-ohm, matched resistors in the cathodes of the 807's produce voltage drops proportional to these currents which are used in an automatic balancing circuit that ensures current equality to within a few tenths of a milliampere throughout the tube life. This "static" balancing circuit which acts to adjust the bias on the 807 grids will be discussed in the next section.

Feedback to the input of the amplifier from the output transformer secondary is adjustable in magnitude and may consist of a combination of negative voltage feedback and positive or negative current feedback. The sign and magnitude of the latter are determined by the 4-ohm potentiometer. In the upper position, switch S_1 removes all current feedback, while in the lowest position it bypasses the 22-ohm resistor with a large electrolytic capacitor, thus making the current feedback effective only at low frequencies.

The addition of variable current feedback as above allows the output impedance of the amplifier to be varied over a rather wide range of positive and negative values.

Combined feedback for output impedance control seems to have first been suggested by Mayer in 1939[13] and has recently been applied in some commercial high-fidelity amplifiers. One of the recent articles on this technique[14] makes the claim that with the internal impedance of the amplifier made sufficiently negative to cancel out the electrical resistance of a loudspeaker load at low frequencies (damping factor -1), better damping of the speaker is obtained than with a very low, positive amplifier output impedance (damping factor 20 to 40). In addition, it is claimed that better transient response is achieved, the low-frequency response of the speaker is extended, its power handling capacity increased, and its low frequency distortion reduced.

The author has been using another amplifier with combined feedback of the above type for more than four years. It can be set to give a damping factor of -1. It must be admitted, however, that with the author's speaker system no significant improvement in low-frequency speaker response can be detected aurally on varying the amplifier impedance from a small positive value to the negative value which gives a damping factor of almost -1. Since the low-frequency speaker of this horn-loaded system has a fundamental resonance of 23.5 cps and negligible harmonic distortion at 30 cps, it may be that there is little or no room for improvement in its low-frequency response. With speakers that are more poorly loaded acoustically, it is possible that some or all of the above claims of the virtues of a damping factor of -1 may be justified to some degree; for this reason, and because of the simplicity of the addition to the circuit, combined feedback was also incorporated in the present amplifier. It is certainly true that an amplifier with a negative output resistance at low frequencies tapering off to a small positive resistance at higher frequencies will tend to increase the bass response of a speaker used with it. It can thus be used to partly or completely correct for a drooping low-frequency speaker characteristic. Such correction must be applied with moderation, however, to avoid driving the speaker outside its range of linear operation at low frequencies.

It will have been noted that three separate supply voltages have been used in the preceding circuits. For this developmental amplifier, all three voltages were derived from electronically stabilized supplies. Since the bias voltage is separately stabilized within the amplifier with a voltage regulator tube, the +250v and −380v supplies do not actually require stabilization. On the other hand, it is very useful to supply the +400v for the output stages from a regulated source. Not only is the output voltage of such a supply easily adjusted, but the low output impedance (0.1-ohm in the author's supply) eliminates the possibility of motorboating and helps to improve the linearity of the output stage for large output signals.

Static Balance Loop

The requirements for automatic static balancing need careful consideration. If the balancing action is too slow in response, the output currents may remain unbalanced for appreciable periods. On the other hand, if the response is too rapid, the "static" balancing circuit will tend to destroy the normal low-frequency push-pull response of the amplifier since it will try to balance out the push-pull signal. It is thus evident that the static balancing circuit is not really static, and a happy-medium response time for this circuit must be selected. In this connection, it might be mentioned that Kiebert[15] has given a low-gain static balancing circuit which apparently has very little higher frequency amplitude discrimination. It may, therefore, be expected to degrade appreciably the low-frequency response of amplifiers with which it is used.

The static balancing circuit used in the present amplifier is shown in Fig. 6 (opposite). Unbalance signals from output tube cathodes first pass through a long-constant RC network necessary to roll off frequency response of static-balance loop above a few tenths of a cycle and to reduce the amplitude of signal-frequency components to such a level that they never overload the static balance amplifier of Fig. 6. The static-balance circuit is itself a self-balancing, direct-coupled differential amplifier which responds only to the difference between the dc (or ultra-low frequency) levels of the output-tube cathode voltages. The 12BZ7 tube of Fig. 6 is connected in a dynamic balancing circuit similar to those of Figs. 2 and 3. It ensures that the

[13] H. F. Mayer, "Control of the effective internal impedance of amplifiers by means of feedback," Proc. IRE, vol. 27, pp. 213–217; March, 1939.

[14] C. A. Wilkins, "Variable damping factor control," *Audio*, vol. 38, pp. 31–33, 66; September, 1954.

[15] M. V. Kiebert, Jr., "System design factors for audio amplifiers," 1954 IRE Convention Record, Part 6, "Audio and Ultrasonics," pp. 25–40 (Fig. 11).

output of the differential amplifier will be truly push-pull and, in addition, gives stabilization of the absolute average output level.

The gain of the static-balance differential amplifier is further stabilized by 20 db of balanced negative feedback from the output back to the input cathodes. The differential gain with feedback is 200 from one input to one output. The common-mode response of this amplifier is exceedingly low. Thus, common changes in the absolute levels of the signals at the two cathodes have no appreciable effect on the output level. The outputs are returned through frequency-sensitive networks to the grids of the driver-amplifier tubes of Fig. 3. It thus becomes clear that the driver-amplifier tubes, the drivers, the output tubes, and the static-balance differential amplifier form a direct-coupled, closed feedback loop of very high gain at ultra-low frequencies.

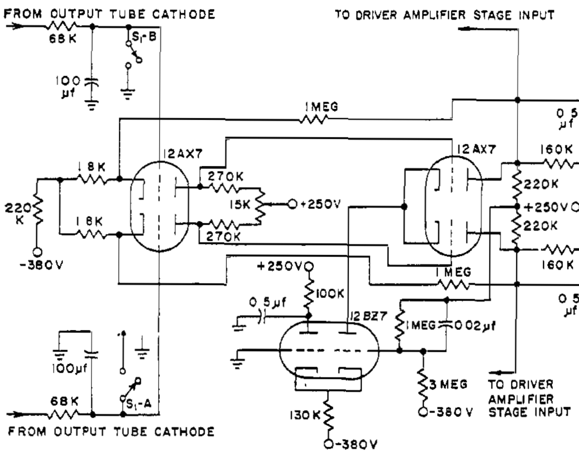

Fig. 6—Static balancing circuit.

With the loop open, measurements show that the limiting low-frequency open-circuit loop gain is 48 db. This is essentially the factor by which differences in the quiescent cathode currents of the output tubes will be reduced by the differential feedback after closing the loop. The switch S_1 of Fig. 6 affords a convenient means of opening the loop without affecting normal operation of the main amplifier. This switch may be used during initial adjustment of the amplifier. After the amplifier has warmed up, the loop is opened and the output tube cathode currents are adjusted to equality by means of the 15K potentiometer in the plate circuit of the first 12AX7 of the static balance circuit. Then after the loop is closed, the feedback has less work to do and can therefore hold the two currents closer to equality. In practice, such an adjustment need never be repeated during the life of the output tubes. It is of interest to note that if one of the output tubes is removed during normal amplifier operation (simulating total failure), the static balance loop entirely cuts off the cathode current of the remaining tube. Unlike an ordinary push-pull circuit which gives a degraded output under such conditions, the present amplifier then produces no output—a much more positive indication of the desirability of tube replacement.

The frequency response of the above feedback loop requires further consideration. The usual Nyquist stability criterion applies to the loop and minimally requires that the upper-frequency open-circuit response of the loop be reduced from 48 db gain to unity gain before a phase-shift of 180 degrees is reached. This condition restricts the rate at which the upper-frequency response can fall with increasing frequency to an absolute maximum of 12 db/octave near unity gain. Unfortunately, this is not yet the whole story. The ordinary amplifier signal is injected into the loop through the 0.5 µf coupling capacitors of Fig. 2. If the present differential-balance feedback loop is temporarily considered to be the main feedback loop, then the signal injected at the 0.5 µf capacitors represents an additional feedback voltage derived from a subsidiary feedback loop which includes the output transformer.[16] At the point of addition of the main and subsidiary feedback voltages, the main feedback voltage (from the balancing loop) decreases with increasing frequency while that from the subsidiary loop increases from zero at zero frequency to a final limiting value at relatively low frequencies. The crossover point of the two voltages essentially determines both the lowest frequency of operation of the amplifier and the response time of the static balancing circuit. In addition, the presence of this subsidiary feedback loop restricts the allowable rate of fall of the main loop response to about 6 db/octave near the crossover region. A faster roll-off causes instability when both loops are closed. This restriction, in turn, requires that the roll-off of the static balance loop begin at ultra-low frequencies so that the crossover point occurs at a low enough frequency that the low-frequency response of the amplifier itself is not appreciably reduced. The 6.8-second RC time constant at the input of the differential amplifier of Fig. 6 yields such a roll-off.

The time constant formed by each 0.5 µf coupling capacitor at the input to the driver-amplifier tubes and the large resistors between each of these grids and the output of the differential balancing amplifier can cause a further roll-off with increasing frequency at a 6 db/octave rate. Since a maximum of only about 6 db/octave in the loop response can be tolerated, this time constant could be used and that at the input of the differential amplifier eliminated were it not for the fact that then the normal amplifier signal appearing at the output-tube cathodes would grossly overdrive the differential amplifier. Because the long input time constant is therefore necessary, the roll-off arising from the 0.5 µf coupling capacitors must be removed. This is accomplished by shunting the 3.3 megohms resistors with the 0.1 µf capacitors shown and isolating the com-

[16] Subsidiary feedback has been discussed by W. T. Duerdoth, "Some considerations in the design of negative-feedback amplifiers," *Proc. IEE*, part III, vol. 97, pp. 138–158; May, 1950.

bination with 100K resistors. The additional 0.5 μf capacitors at the outputs of the differential amplifier become effective in further rolling off the response only at sufficiently high frequencies (greater than 5 cps) that the gain of the balancing loop is less than unity. This fairly complicated tailoring of the asymptotic gain characteristic of the balancing loop results in complete stability, sufficiently fast response time, and excellent low frequency response of the main amplifier itself. The balancing loop has no effect on the transient behavior of the main amplifier so long as signal components below about 1 cps are avoided.

We have now completed discussion of all parts of the main amplifier except feedback. The complete block diagram of Fig. 7 shows all the loops except that for optional positive voltage feedback which will be discussed later. The letters *DB* and *SB* in this diagram stand for dynamic balance and static balance. It will be seen that there are, altogether, ten loops in the amplifier. Only those marked α and β are effective in changing amplifier gain; the others all improve its performance in the various other ways which we have already discussed. Note that the combination of negative voltage and positive or negative current feedback from the output is only symbolically indicated in Fig. 7; the actual circuit employed is given in Fig. 5.

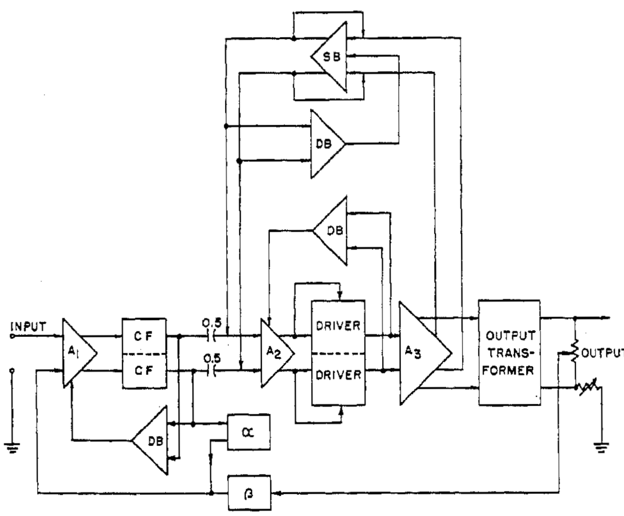

Fig. 7—Complete block diagram of the amplifier.

Feedback Loops to the Input

The main function of feedback in an audio amplifier is to reduce nonlinear distortion. Other desirable results of feedback are a reduction in amplifier output impedance and flattening of the gain-frequency characteristic in the working range. Fig. 8 shows the high frequency gain characteristics of the present amplifier with no over-all feedback at various stages.[17] For easy comparison, all three curves have been normalized to have the same value at low frequencies. The output frequency response is down by 1 db at 38 kc; it is therefore apparent that the response is adequate for audio applications even without feedback. On the other hand, feedback is desirable to reduce the nonlinear distortion of the amplifier, particularly when it is operated in class B.

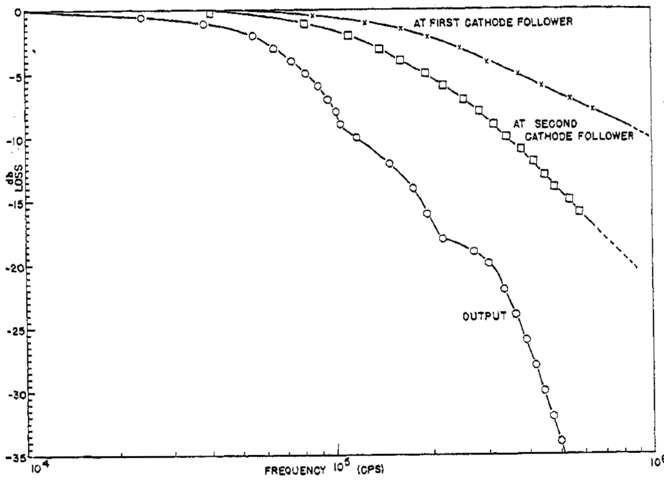

Fig. 8—Gain-frequency characteristics at various points in the amplifier with no over-all feedback.

Feedback is applied in an amplifier in order to make the output signal as close an amplified replica of the input signal as possible. The larger the fraction of the output which is compared with the input, the nearer the desired condition of exact similarity will be approached. It is therefore obvious that any local negative feedback loops within an amplifier which reduce its over-all gain will restrict the amount of output-input negative feedback which may be applied for a given final amplifier gain. For this reason, it is desirable that such local loops be avoided whenever possible so that maximum output-input negative feedback may be employed. Such a system necessarily gives maximum nonlinear distortion reduction. This conclusion has been pointed out before[12,18] but it bears repetition in view of the appearance of recent amplifier articles which stress the advantages of local loops.[10,11] One of these articles,[11] in justification for its point of view, states that, "This approach (the use of output-input feedback) in a multi-stage amplifier, results in a peaked response at both ends of the audio spectrum and a narrow margin of stability." As we shall see, neither of these conclusions is necessarily correct.

For the above reasons, it was decided to use as much over-all negative feedback in the present amplifier as possible, consistent with a final gain of about 30 db. Since the gain of the amplifier with no feedback and with a 15-ohm load was 61 db, some 30 or 31 db of feedback could be applied and still leave sufficient gain. It

[17] Unless otherwise stated, all measurements shown in the figures were taken with an average output tube bias of -45v and with a 15-ohm load resistor. For the frequency response measurements of the present section, the output power level was about 1 watt.

[18] W. B. Bernard (Letters), *Electronics*, vol. 27, pp. 401–404; January, 1954; pp. 372–376; December, 1954.

was first found that using a single feedback resistor shunted with a small capacitor, up to 18 db of over-all negative feedback could be applied before small parasitic oscillations appeared in the output at high power levels. This value is very consistent with the output response curve of Fig. 8.

To ensure stability in a feedback amplifier, it is necessary that the complex feedback factor βG not enclose the point $-1 \measuredangle 0$ on a polar plot. Here G is the complex gain of the amplifier without feedback but loaded with its usual load of 15 ohms. For convenience, we have taken the midfrequency value of βG to be positive real when it represents negative feedback, and shall use the same sign convention for other feedback factors. Further, to eliminate the peaked response at the ends of the audio spectrum mentioned earlier, it is necessary that the absolute value of the gain with feedback G', given by $G/(1+\beta G)$, not exceed its midfrequency value. In addition, it is desirable that even after G' has decreased appreciably from its midfrequency value, there be no secondary rises of G', even though such peaks do not reach the midfrequency value. Exact specification of the form of βG which meets the above requirements is complicated; in practical cases, it is usually only necessary to ensure a wide stability margin such as that produced by restricting the phase angle θ of βG to the range -90 degrees $\leqq \theta \leqq +90$ degrees, as long as $|\beta G|$ is greater than or equal to unity. This condition requires that the roll-off of βG at either end of the spectrum does not appreciably exceed a rate of 6 db/octave until after $|\beta G|$ is less than unity.

The stability conditions can be met by changing and controlling the frequency response of β, of G, or by a combination of these methods. An additional method is that of subsidiary feedback discussed by Duerdoth.[16] If a negative feedback voltage derived from the output of the first stage is added to βG at the input,[19] the combination can be made to have characteristics yielding excellent stability. If this subsidiary feedback were effective at all frequencies, it would be of the undesirable "local-loop" feedback type already discussed. However, if its magnitude is greatly reduced in the working band of the amplifier and is only greater than unity (referred to the input signal amplitude) beyond either or both ends of the working band, then it will not appreciably affect conditions within the working band. Since the subsidiary feedback voltage is derived after only a single stage of amplification, it will automatically have a high-frequency limiting slope of only 6 db/octave. At very high frequencies, it will decrease less rapidly than βG and will, therefore, eventually dominate the sum of αA_1 and βG. This sum, the over-all feedback voltage, therefore will have the desirable limiting decay rate of 6 db/octave.

[19] We shall use the terms feedback factor (such as βG) and feedback voltage interchangeably for convenience. The actual feedback voltage of an opened loop corresponding, e.g., to βG is βG times the amplifier input voltage.

The combination of over-all and subsidiary feedback gives a final amplifier gain G' of $G'=G/(1+\alpha A_1+\beta G)$. Calculation shows, however, that the factor by which nonlinear distortion arising in the second or third stages is reduced by the combination is not $(1+\alpha A_1+\beta G)$ but $(1+\alpha A_1+\beta G)/(1+\alpha A_1)$ instead. It is therefore particularly desirable to ensure that $|1+\alpha A_1|$ is not appreciably greater than unity within the working band.

The present amplifier makes use of high-frequency subsidiary feedback of the type discussed above to allow 30 db or more of feedback to be used with a wide stability margin. The subsidiary feedback voltage is taken from point a of Fig. 2, passed through the 200 $\mu\mu$f capacitor and 1K resistor shown and added to the over-all feedback. The adding takes place across the series combination of the 1K resistor, 1.5 mh choke, and 100 μf capacitor shown in Fig. 5. The 200 $\mu\mu$f capacitor and 1K resistor in the subsidiary path ensure that the subsidiary feedback factor αA_1 is negligible in the working band.

The main over-all feedback is adjustable by the 2K potentiometer of Fig. 5. It is further reduced about 20 db at low and medium frequencies by the 10K series resistor. The phase angle of the over-all feedback is improved at high frequencies both by the small capacitors in parallel with the 10K resistor and by the 1.5 mh choke. Subsidiary feedback is unnecessary for stabilization of the amplifier at very low frequencies. An adequate stability margin is produced by the 100 μf capacitor of Fig. 5 in series with the 1.5 mh choke. At very low frequencies, it reduces the roll-off rate of βG to an acceptable value near 6 db/octave. It is found that with the present combination of subsidiary and over-all feedback, more than 40 db of over-all negative feedback may be applied in the working band of the amplifier with complete stability.

Results of the above stabilization technique are shown in Figs. 9 and 10 (next page) measured with 30 db of midfrequency negative feedback. Fig. 9 shows G and G' vs frequency plotted on an absolute gain scale. It is seen that the limiting decay rate of G is 24 db/octave. Nevertheless, the amplifier is stable and G' shows no peaks. With feedback, the amplifier is flat to within 1 db from below 5 cps to 72 kcps. Fig. 10 shows how the various feedback voltages add at high frequencies. It also is plotted to show both absolute and relative amplitudes. Measurements were not extended to sufficiently high frequencies to show that the combined characteristic $(\beta G+\alpha A_1)$ has a final limiting slope of 6 db/octave, but such behavior must be finally reached. The curve including the factor $-\rho A_1$ will be discussed later.

The magnitude of the feedback effective in reducing distortion is of interest. Since it is given by the quotient of two complex terms, it cannot be inferred directly from measurements which do not include phase angles. However, up to 50 kc or so, the phase angle of $(1+\alpha A_1$

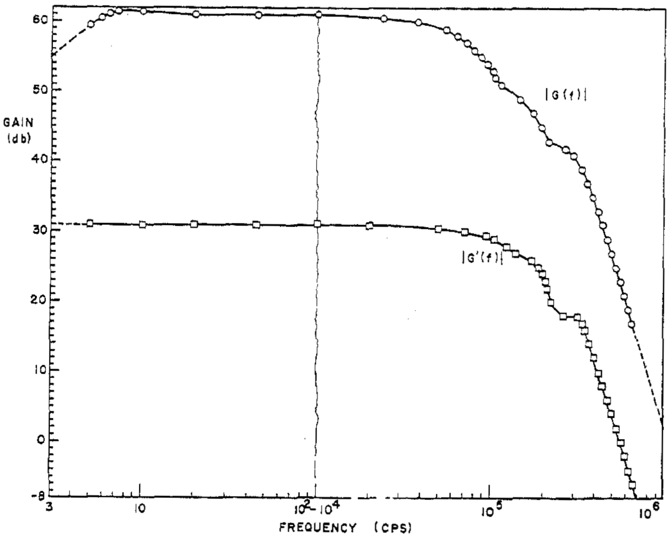

Fig. 9—Loaded amplifier gains $|G|$ (without feedback), and $|G'|$ (with 30 db of over-all feedback) *vs* frequency.

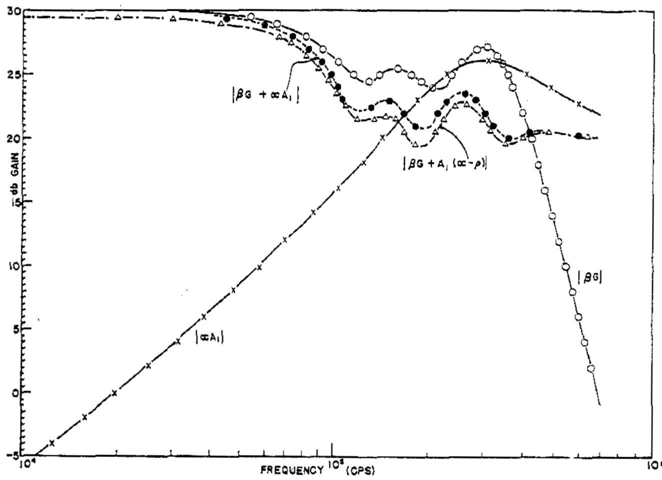

Fig. 10—High-frequency variation of the amplitudes of various feedback factors.

$+\beta G)$ is very close to zero. Hence, we need be concerned only with the magnitude of $|1+\alpha A_1|$. Again up to 50kc or more, it is obvious from the curve of $|\alpha A_1|$ of Fig. 9 that the phase angle of αA_1 must be close to 90 degrees. Its magnitude at 20kc is zero db, or unity. Hence $|1+\alpha A_1|$ must be $\sqrt{2}$ at 20kc, or 3 db. Similarly, it is found that this quantity is 1 db at 10.8kc and about 6 db at 35kc. When these factors are subtracted from the midfrequency feedback of 30 db, one obtains the resultant feedback effective in reducing distortion at the given frequency. For example, this so-called harmonic feedback is 27 db at 20kc.

The above results show that the addition of subsidiary feedback has reduced the effective feedback by only an unimportant factor within the working band of the amplifier to 20kc. In order to keep the harmonic feedback nearer 30 db out to frequencies of the order of 70 to 100kc, the low-frequency roll-off of the subsidiary feedback can be made more rapid than the present 6 db/octave. Using a constant-k filter in the subsidiary loop, such a result was indeed obtained. It was found that although it did result in the maintenance of increased negative feedback to much higher frequencies, it also produced an initial peak in the high-frequency square-wave response. More careful control of the characteristics of the two feedback voltages in the neighborhood of their crossover undoubtedly could eliminate this effect, but the extra effort necessary was felt unwarranted for the present amplifier.

The addition of the over-all and subsidiary feedback voltages has been carried out so as to minimize $|1+\alpha A_1|$ in the working band and to produce the best possible high-frequency square-wave response. The small, variable 50 $\mu\mu$f capacitor across the feedback resistor of Fig. 5 is used to adjust for best response. With the best adjustment, 10kc square waves show almost square corners and flat tops and bottoms with at most only a trace of greatly damped high-frequency parasitic oscillation on the tops. The rise time of the amplifier with 30 db over-all feedback and subsidiary feedback is 3 μs between 10 and 90 per cent amplitude points. The recovery of the amplifier from overloads also occurs in approximately this interval. As the output square-wave amplitude is increased, the only change in square-wave shape is a slight sharpening of the upper left corner. More than 0.01 μf of external capacitance can be connected in parallel with the load without any appreciable effect on the shape of square waves. Further, square-wave shape and amplitude are virtually independent of load from open circuit to a very heavy load. It was also found, using a direct-coupled oscilloscope, that the tilt of low-frequency square waves of any amplitude within the amplifier capabilities was 7, 9, and 10 per cent respectively, at 30, 20, and 17 cps. These results are consistent with a measured phase angle of -8 degrees at 5 cps.

Although the above over-all harmonic feedback of almost 30 db over the entire working band of the amplifier results in exceptionally low output distortion, it was of interest to see if this distortion could not be reduced even further by the addition of a positive voltage feedback loop around the first stage in the manner of Miller.[20] If the positive feedback factor is denoted by $-\rho A_1$, then the combination of over-all, subsidiary and positive feedback results in an output gain reduction factor of $N=(1+\beta G+\alpha A_1-\rho A_1)$. Nonlinear distortion in the first stage is also reduced by this factor. Distortion produced in the rest of the amplifier is reduced by the factor $M=N/[1+A_1(\alpha-\rho)]$, however. In the working band, the factor αA_1, may be neglected. We see that if we make ρA_1 equal to unity in the working band, the factor $[1+A_1(\alpha-\rho)]$ approaches zero and M becomes very large. Thus, a small amount of positive feedback

[20] J. M. Miller, Jr., "Combining positive and negative feedback," *Electronics*, vol. 23, pp. 106–109; March, 1950.

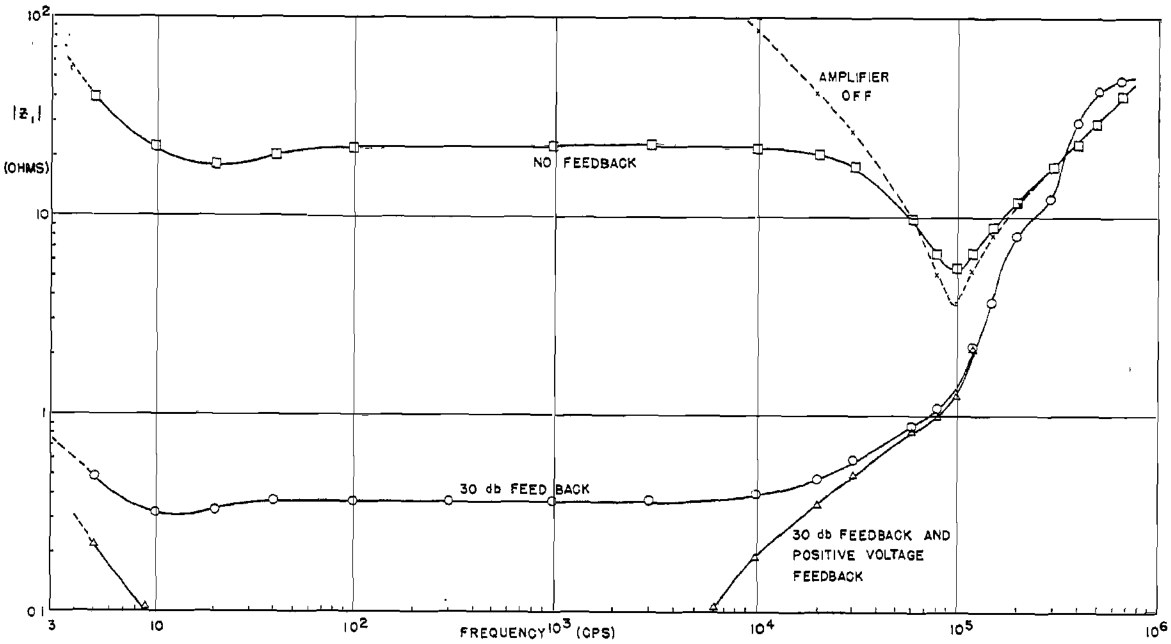

Fig. 11—Frequency dependence of the magnitude of amplifier internal impedance for various feedback conditions.

can reduce the output distortion of the amplifier by a large factor. Although the distortion of the first stage is increased slightly by the addition of positive feedback, it can be made so low initially that a small increase is immaterial. Since the maximum distortion of usual amplifiers occurs in the last stages, it is particularly helpful that the distortion reduction of positive feedback is most effective for these stages.

Positive feedback is taken from the point b of Fig. 2. It passes through a 1 μf capacitor, an on-off switch S_1, and the 25K potentiometer and 43K fixed resistor. It is then added to the other feedback voltages and is applied to the lower input grid of Fig. 2. The 470K resistor connected to the off position of S_1 is used to keep the coupling capacitor charged so that switching transients are eliminated on turning the positive feedback on and off. The combination of positive and negative feedback will be stable so long as the composite feedback factor $[\beta G+A_1(\alpha-\rho)]$ does not enclose the $-1\measuredangle 0$ point on a polar plot. It is therefore necessary that the positive feedback decrease more rapidly outside the working band (or change over to negative feedback) than does the negative feedback. In the present amplifier, stability is ensured at the low-frequency end by the 1 μf capacitor which reduces ρA_1 to zero at zero frequency and shifts its phase by 90 degrees. At the same time, this capacitor is still sufficiently large that positive feedback is fully effective down to the lowest signal frequencies to be passed by the amplifier. At the high-frequency end, the progressive increase in αA_1, a negative feedback factor, soon cancels out the positive feedback beyond the working band. The 25K potentiometer shown is used to adjust ρA_1 to exactly unity for midfrequencies. Fig. 10 shows the effect of $-\rho A_1$ on the composite feedback factor

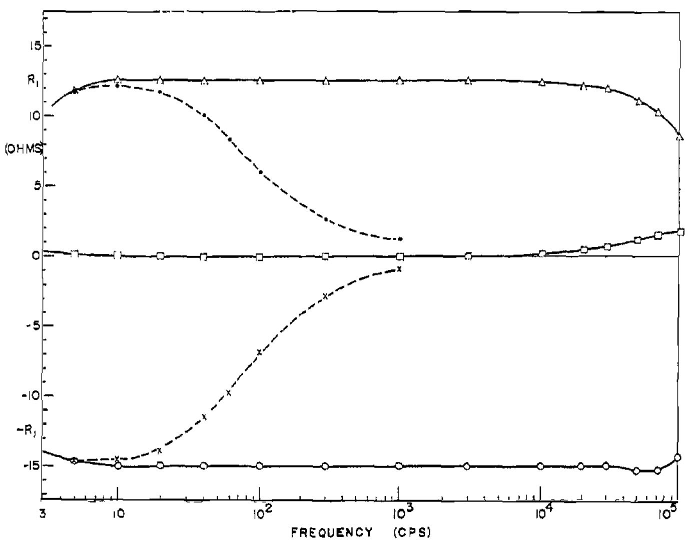

Fig. 12—Frequency dependence of amplifier internal resistance with 30 db over-all negative voltage feedback and positive or negative output current feedback.

at high frequencies. For this measurement, ρA_1 was adjusted to unity in the working band.

An excellent indirect measure of the effectiveness of feedback in reducing nonlinear distortion is afforded by its reduction of amplifier output impedance. We shall discuss distortion results later; meanwhile, Figs. 11 and 12 present measurements of amplifier internal impedance $|Z_i|$ for various feedback conditions. These measurements were made with a bias of -41 volts. With no feedback, the magnitude of this impedance is about 22 ohms over the working band, slightly larger than the nominal value of 15 ohms. The internal impedance of the amplifier without feedback increases with the magnitude of the bias value used. The above internal impedance value indicates that an output-transformer primary winding impedance somewhat greater than the 2,500 ohms employed might be desirable for best operation. However, the present transformer impedance ratio, 2,500 to 15, is close to the ideal in terms of low distortion and maximum output power, as we shall see in the next section. The application of 30 db of over-all feedback reduces the impedance to about 0.36 ohms over most of the working band. Finally, the application of positive voltage feedback reduces $|Z_i|$ to a value no greater than 0.01 ohms over much of the working band. Phase measurements indicated that Z_i was essentially resistive over this band.

Fig. 12 shows the output impedance with no positive voltage feedback but with negative or positive current feedback from the output, in addition to the usual 30 db over-all negative voltage feedback. The top and bottom curves are the results for the two extreme settings of the 4-ohm potentiometer of Fig. 5. They are actually presented as positive and negative resistances since phase measurements indicated that the output impedance was almost entirely resistive over the greater part of the band shown. The two dotted curves are similar to the outer curves, but the 100 μf bypass capacitor of Fig. 5 is used to limit the output current feedback to low frequencies. The dotted curves are not extended beyond 10^3 cps since the output impedance has appreciable phase shift, arising from the bypass capacitor, above 10^2 cps for either positive or negative current feedback. The output impedance for either condition eventually reaches the small positive value shown in Fig. 11 for negative voltage feedback alone. The center curve is the output impedance or resistance measured with the 4-ohm potentiometer adjusted for zero impedance in the midfrequency range. We see that it remains essentially zero over a wide frequency band.

It is of interest to give the expression for Z_i when negative voltage feedback βG, negative subsidiary feedback αA_1, positive voltage feedback $-\rho A_1$, and positive or negative output current feedback may all be present simultaneously. Neglecting loading of the various feedback paths by each other (a good approximation in the working band), we find for the present amplifier

$$Z_i = \frac{[1 + A_1(\alpha - \rho)][Z_0 + R_+ + R_-] + [1 + (Z_0 + R_+ + R_-)/Z_L][(\gamma + \beta)R_+ + (\gamma - 1)R_-]G}{1 + A_1(\alpha - \rho) + [1 + (Z_0 + R_+ + R_-)/Z_L]\beta G},$$

where Z_0 is the amplifier internal impedance without feedback. The current feedback resistors R_+ and R_- are indicated in Fig. 5; the quantity γ specifies the setting of the 4-ohm potentiometer of this figure. We see from the above formula that Z_i may be made zero by adjustment of either the positive voltage feedback factor $-\rho A_1$ alone, by positive current feedback alone, or with both types simultaneously. The foregoing measurements are in general agreement with the above formula, and the amplifier is completely stable with any combination of the above types of feedback.

When output current feedback in addition to over-all negative voltage feedback is employed, the effective gain with feedback, G', is given by

$$G' = \frac{G}{1 + A_1(\alpha - \rho) + [\beta + (\gamma + \beta)R_+/Z_L + (\gamma - 1)R_-/Z_L]G}.$$

Thus, negative current feedback decreases G' and positive current feedback increases it. We have dealt throughout thus far with the loaded gains G and G' since it is these gains which are important in practice. For some purposes, it may be desirable to express the above formulas in terms of the unloaded gain K of the amplifier. The relation between G and K is $G = K/[1 + (Z_0 + R_+ + R_-)/Z_L]$. If usual amount of feedback employed in present amplifier were expressed as the difference between, unloaded gain with and without feedback, rather than between loaded gains, instead of 30 db it would amount to between 36 and 38 db.

Power and Distortion Measurements

Fig. 13 (opposite) shows maximum "undistorted" power output of amplifier vs frequency. This measurement was made using a Ballantine type 310A ac voltmeter to indicate rms load voltage and an oscilloscope to indicate distortion. Power output of amplifier was increased at a given frequency until distortion could be observed on the oscilloscope then reduced to the point where distortion was no longer detectable. From comparison with intermodulation measurements it is estimated that the actual distortion level of the "undistorted" power output shown in this graph probably does not exceed 2 per cent intermodulation and is less over much of the range.

It may be calculated from Fig. 13 that the maximum output power is flat within minus 0.5 db from 19.8 cps to 22.4 kcps. The roll-off on the low-frequency end has a slope of about 6 db/octave and is due to the inability of the output tubes to supply sufficient magnetizing current to the transformer at very low frequencies. On the other hand, the high-frequency roll-off has an initial slope of 3 db/octave which probably arises from the inability of the output tubes to supply the necessary charging currents for the transformer primary winding capacitance. In the midfrequency region, the maximum power is limited by the diode line of the output tubes, the point where the grids lose control of the plate currents of the output tubes. The drivers of the amplifier can drive the grids considerably more positive than the diode line with negligible distortion, but as soon as the diode line is reached and exceeded, the output signal begins to show symmetrical peak clipping.

The point where the diode line is intersected by the load line of the output tubes is determined (for fixed load resistance) by the plate supply voltage available. Increasing this voltage increases the maximum power available before diode-line clipping becomes apparent; it was found that with an E_{bb} voltage of the order of 500v, more than 80w of output power could be obtained before such clipping occurred. Such supply voltages of course exceed the rating of the 807 output tubes.

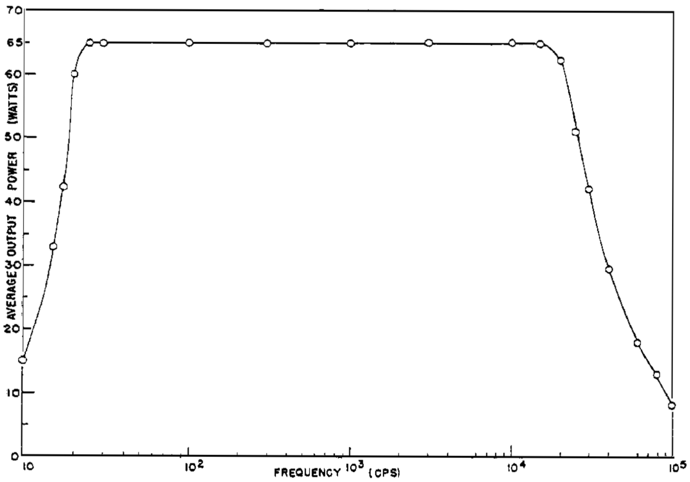

Fig. 13—Maximum "undistorted" average power output *vs* frequency.

In audio applications, very high power peaks occur quite seldom; yet when they do, it is desirable for the amplifier to handle them without clipping if possible. It occurred to the author that since an electronically stabilized power supply for E_{bb} was used with the present amplifier, it might be possible to drive the power supply with the driver signal from the amplifier in such a way that above 60w or 65w output, the value of E_{bb} increased automatically with the signal level from its usual value of +400v to whatever higher value was required to eliminate diode-line clipping at the given signal level. Such an increase in E_{bb} is eventually limited by the quiescent voltage drop initially available across the series current tube of the electronically stabilized supply; an increase of more than 100v above +400v could be obtained in the author's supply. The above idea was investigated in a preliminary fashion using biased diode rectifiers to derive a power-supply driving signal from the amplifier drivers only when the driver signal exceeded the value which resulted in an output power of about 65w. Quite an appreciable increase in unclipped output power could be obtained in this fashion. The circuit is not described in detail here since output powers exceeding 65w were not really necessary for the author's application. It should be noted that although the maximum rated value of dc plate voltage is exceeded in the above scheme, this may be expected to occur only rarely in most applications. This operation of the output tubes, therefore, should be satisfactory in terms of tube life. Should greater output power than 65w be required without the use of the above scheme, tubes with higher ratings such as the new type 6550 could be used in place of the 807's. It may be mentioned that this idea of driving a regulated power supply may be used with a driving signal from the amplifier output arranged to reduce the supply voltage after a given delay in order to allow the amplifier to produce full power for short periods but only reduced power for longer periods, thus avoiding loudspeaker damage. A similar idea is presently employed in a high-power commercial amplifier. The use of a fast-acting fuse in the speaker line will achieve the same result in a much simpler fashion, however.

Triode-connected, push-pull 807 tubes operated in class AB_1 are rated at only 15w output power. Operating them in class AB_2, B_2, or A_2 as in the present amplifier results in more than a fourfold increase in output power. We see that the following quoted statements are therefore unjustified: "Obtaining 15 to 20 watts output without using four output tubes and a large power supply, and without operating the power tubes beyond ratings, rules out a triode output stage;"[10]—"push-pull class *A* 6L6's will give 18.5 watts output with two-percent harmonic distortion. There are no receiving-type triodes which will match this performance . . . there is no longer any reason to use triodes in amplifiers up to 100 watts."[18] Although the 807 is not strictly a receiving-type triode, comparable results to those of the present amplifier could be obtained using 6L6's, 5881's, or 6AR6's triode-connected, operated within their ratings up to 30 to 50w output. For the same or greater power output, the advantages of triodes over beam-power tetrodes and pentodes in terms of lower output impedance and less higher-order harmonic distortion scarcely have to be mentioned.[21]

The noise level of the present amplifier with no input signal and the usual 30 db of over-all feedback was found with the Ballantine 310A meter to be about 95 db below 60w output. In spite of the use of ac-operated tube heaters with one side grounded and the use of a number of cathode-followers, no hum signal could be distinguished in the output. This output consisted largely of unavoidable radio signals picked up on the measurement leads. When these signals were filtered out at the voltmeter, the residual noise voltage was about 106 db below 60w and was equivalent to an input signal of about 5 μv rms. A single ground buss connected to chassis at input is used in this amplifier.

The power efficiency of the amplifier is of some interest. This efficiency is greatest for class B operation and since, as we shall see later, distortion is acceptably low even for this mode of operation, we shall consider its

[21] F. Langford-Smith, "Radiotron Designer's Handbook," Amalgamated Wireless Valve Co., Ltd., Sydney, Australia, 4th ed., pp. 546 *et seq.;* 1952.

efficiency only. At full-power output of 65w, the total plate input is about 125w, slightly exceeding the rated value of 100w. The plate efficiency is, therefore, about 52 per cent and the plate dissipation per tube 30w, the maximum ICAS rating. The total efficiency should include the grid power input, but this is difficult to estimate because of the peaked shape of the grid current waveform. Measurement indicates that it certainly does not exceed 8w at maximum power output; thus, it does not change the efficiency very much. Of the order of 2w of power are dissipated in the winding resistances of the transformer (about 90 ohms total referred to full primary) at full-power output. About another half-watt is dissipated in the 2K feedback potentiometer across the output.

The above results were obtained with average grid bias of about −55v; the quiescent plate currents under these conditions were 6 milliamperes each, giving a quiescent plate dissipation of 2.4w for each tube. This low value is a good indicator of long tube life; very seldom will the full 65w of power be required from the amplifier under ordinary conditions. About 60w of power can be obtained continuously from the amplifier, however, without even exceeding the CCS ratings of the 807's. By increasing the load resistance seen by the output tubes, the plate efficiency can be appreciably increased if desirable and the distortion reduced; at the same time, however, the maximum available power is reduced.

A considerable amount of nonlinear distortion measurements on the present amplifier has been carried out by various methods. We shall first discuss the results of intermodulation (IM) measurements with the usual SMPE rms sum method.[22] Because of the very low distortion of the amplifier under certain conditions, a commercial intermodulation tester was modified to reduce its own nonlinearity and to improve its filtering and was then accurately calibrated from 30 per cent to 0.04 per cent IM.[23] Intermodulation readings were made on the Ballantine 310A rather than the meter of the test set. In spite of these precautions, the readings obtained in the neighborhood of 0.05 per cent may be slightly high. For the following measurements, output power was found using the General Radio Type 783-A output power meter set to 15 ohms load.

The intermodulation distortion of the amplifier with a bias of −45v and no subsidiary or over-all feedback is presented in Fig. 14 as a function of equivalent single-frequency average output power, This bias value corresponds to class AB operation. Greater distortion over most of the range was observed with class B, less with class A. In all the SMPE IM distortion measurements, essentially equivalent results were obtained using either 60 and 5,600 cps or 60 and 2,500 cps signals mixed so

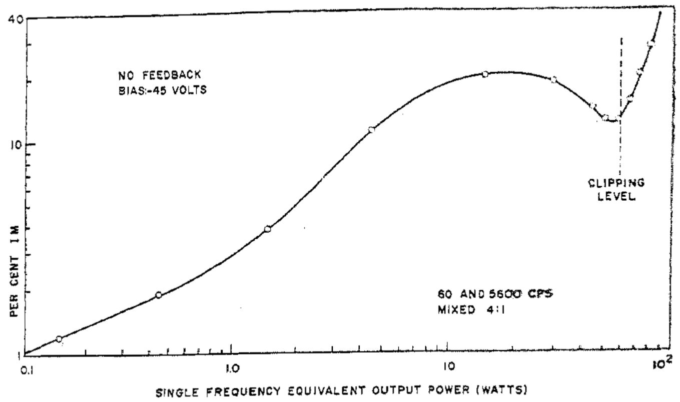

Fig. 14—Percentage intermodulation distortion *vs* output power; no feedback, class AB.

that the 60-cps signal was four times greater than the higher frequency signal.

Figs. 15, 16, and 17 show the IM results for class B, AB, and A with 30 db of over-all feedback and without and with the local positive voltage feedback already discussed. Even without the positive feedback, the amplifier shows an extremely low distortion for all three biases right up to the peak clipping point (about 65w). With positive feedback, the distortion is even smaller. We see that, in general, the larger the distortion obtained without positive feedback, the more effect the positive feedback has. The positive feedback adjustment was unchanged during measurements for a given bias but was adjusted slightly differently for the different biases in order to give minimum distortion over the whole power range for each curve. Positive feedback can do little to reduce distortion once peak clipping commences and the final steep rise of distortion begins.

Some of the pertinent results shown in Figs. 15 through 17 are summarized in Table I which gives the output power at 0.1 and 1 per cent IM distortion for the

TABLE I

OUTPUT POWER (WATTS) AT TWO INTERMODULATION DISTORTION LEVELS FOR VARIOUS BIAS AND FEEDBACK CONDITIONS

Per cent IM	Bias = −57 v Class B		Bias = −45 v Class AB		Bias = −35 v Class A	
	No Positive Feedback	Positive Feedback	No Positive Feedback	Positive Feedback	No Positive Feedback	Positive Feedback
0.1	—	6	1.5	45	20	33
1.0	0.8, 37 63	65	66	67	62	63

various bias and feedback conditions. We see that, as expected, class A operation is superior at the lower power levels. At the maximum power levels just before and after peak limiting begins, it is very interesting to note, however, that the −45v bias gives the best result at a given power or distortion level. It was expected that class B operation, with its greater bias, would be superior in this region. Although the differ-

[22] *Ibid.*, pp. 612–613.

[23] J. R. Macdonald, "The calibration of amplitude modulation meters with a heterodyne signal," PROC. IRE, vol. 42, pp. 1515–1518; October, 1954.

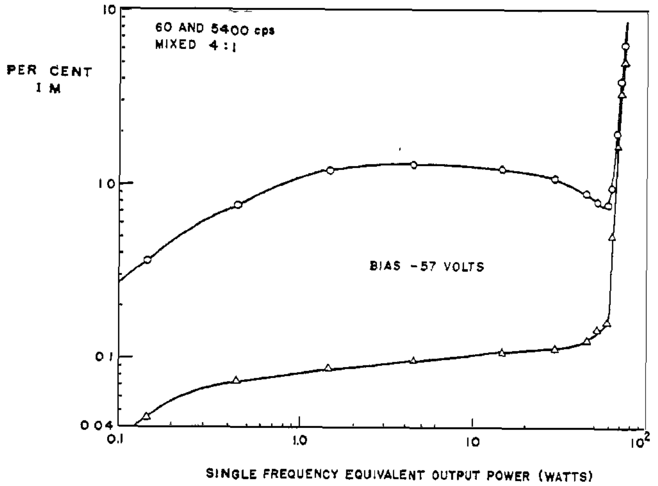

Fig. 15—Percentage intermodulation distortion *vs* output power; 30 db over-all feedback, class B. Lower curve taken with positive voltage feedback.

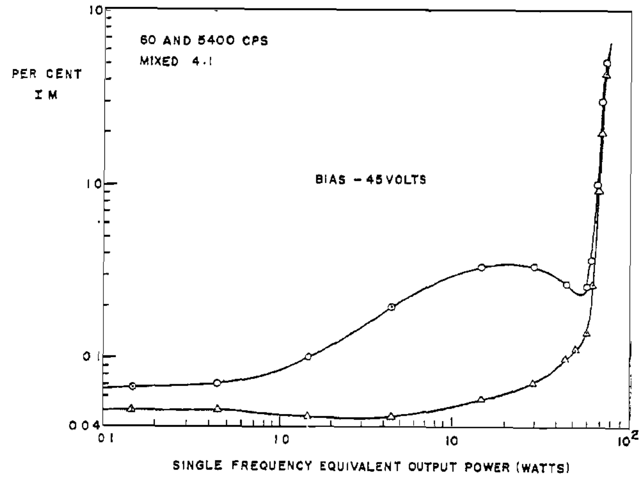

Fig. 16—Percentage intermodulation distortion *vs* output power; 30 db over-all feedback, class AB. Lower curve taken with positive voltage feedback.

ence between the results with -45v and -57v bias is slight in this region, it is still large enough to be significant. Therefore, automatic bias control, which would increase the bias at high signal levels, is undesirable here, and a fixed bias of about -45v gives the maximum power-over-distortion quotient at high levels.

For most practical purposes, the differences between the curves obtained with class B and Class AB operation are only of academic interest, and the amplifier can be operated in class B for minimum quiescent power, maximum plate efficiency, and longest tube life. Even the largest value of IM distortion obtained with class B operation without positive feedback (1.3 per cent) is much too small to be audible. Measurements of the amplifier linearity characteristic (output vs input) with or without positive voltage feedback show no measurable deviations from linearity over the useful dynamic range of the amplifier limited at the low end by its intrinsic noise output of about 2 millimicrowatts and at the high end by diode-line peak clipping.

Several other intermodulation measurements of the above type have been made on the amplifier. In particular, it is found that IM distortion at the driver outputs with output tube loads is less than 1 per cent at a signal level giving maximum power output, when no over-all feedback is employed. Most of the output distortion at all levels thus occurs in the output stage. Measurements of the output distortion as a function of load resistance R_L with 30 db of over-all but no positive feedback, show that it decreases proportional to R_L^{-n} over the range from 5 to 50 ohms with the exponent n slightly greater than unity. An increase in load resistance also increases the plate efficiency of the output stage, but the maximum output power is reduced. A measurement of output power vs load at the 2 per cent IM level showed that maximum power was available with a load of about 10 ohms but the increased power was only slightly greater than that available with the usual 15-ohm load.

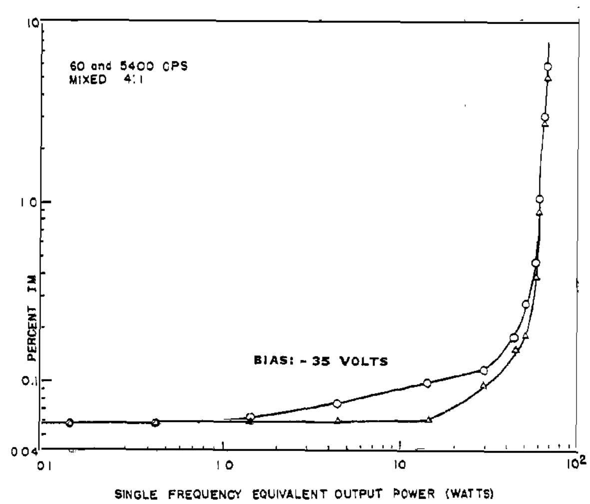

Fig. 17—Percentage intermodulation distortion *vs* output power; 30 db over-all feedback, class A. Lower curve taken with positive voltage feedback.

In addition to the above SMPE intermodulation measurements, it was desirable to check the nonlinear distortion over a wide-frequency band to make sure that it did not rise greatly at the ends of the working band. All the succeeding measurements were made with a bias of -45v. First, a General Radio Type 736-A wave analyzer was used to measure harmonic distortion directly. Because of unavoidable oscillator distortion, it could only be verified that the total harmonic distortion, D_h, was less than 0.2 per cent from 20 to 5,000 cps (the limits of measurement) when 30 db over-all feedback was employed, and positive voltage feedback used or not. The distortion did not exceed the above value at any frequency in this range until clipping began, and a value of only 0.85 per cent was obtained with visible clipping at slightly greater than 65w output power.

Next, harmonic distortion was measured at the 30w level on the amplifier without any feedback. A value of 4.65 per cent was found for D_h from 5,000 to 300 cps. Below 300 cps, the distortion rose slowly to 5.3 per cent at 32 cps and to 5.9 per cent at 20 cps. The distortion was almost entirely made up of third and fifth harmonics with the fifth three times smaller than the third. Since we have shown already that feedback is fully effective down considerably below 20 cps, we can have confidence that with feedback applied D_h is exceedingly low even at the lowest frequencies to be passed by the amplifier.

The CCIF difference-frequency intermodulation method[24] was then used to investigate amplifier behavior at medium and high frequencies. Here two signals of equal amplitude and frequencies f_1 and f_2 are applied to the input, and f_1 and f_2 are varied in such a manner that their difference Δf remains constant. Using a wave analyzer, the distortion component D_d at the difference frequency Δf is measured and expressed as a percentage of either applied voltage. With the usual 30 db of over-all feedback, this measurement showed that D_d was less than 0.2 per cent from 500 to 15,000 cps using a value of Δf of 100. The value 0.2 per cent represented the residual intermodulation in the amplifier input signal arising from oscillator pulling. Without feedback, and using signals of 1,000 and 1,100 cps, D_d was found to be about 0.26 per cent at a 15w level.[25] This result may be compared with the value of D_h of 4.65 per cent and of rms-sum intermodulation of 18.5 per cent at the 30w power level. The ratio of the last two figures is 3.97, satisfactorily close to the value of 3.84 to be expected for third harmonic distortion alone.[26]

Although the residual value of D_d in the input signal applied to the amplifier was too large to allow accurate measurements of D_d to be made when feedback was used, this was not the case with distortion components occurring at frequencies of $(2f_1-f_2)$ and $(2f_2-f_1)$. These two components had approximately equal amplitudes under all conditions, and the ratio of their average value to the rms value of either input signal will be denoted by D_n. Its value for the amplifier without feedback at a 15w average power level was 2.94 per cent. With 30 db of over-all feedback, the value dropped to 0.06 per cent, and with positive voltage feedback applied the value dropped at least another factor of ten to below 0.006 per cent. Note that the peak rms power under these conditions was 60w.[25]

The intermodulation components at $(2f_1-f_2)$ and $(2f_2-f_1)$ are considerably larger than that at (f_2-f_1) and seem to be the largest intermodulation distortion components under most conditions. They arise entirely from third-harmonic type distortion. The fact that they are far below the level of audibility may be graphically illustrated by considering the above results where the rms value of either of the applied voltages was 15v and that of either of the above largest intermodulation components was 9 mv with 30 db of over-all feedback. The average output power with both signals applied is 15w; the output power in either of the above distortion components is about 5 microwatts. With positive voltage feedback in addition, this power drops to less than 50 millimicrowatts. No wonder such distortion products are masked by the undistorted components of the output signal.

Finally, the quantity D_n was measured for values of f_1 between 1 and 15 kcps keeping Δf equal to 100 cps. Measurements were made with 30 db over-all feedback and with and without positive voltage feedback. Without the latter feedback, D_n had increased by only 2 per cent at 15 kcps over its 1,000 cps value. The effectiveness of the positive feedback decreased appreciably with frequency however, and at 15 kcps it only reduced D_n by an additional factor of two. This decrease in effectiveness arises from the increase in the subsidiary feedback factor αA_1 with frequency. By keeping the latter quantity smaller out to higher frequencies, positive voltage feedback could be made even more effective in the high frequency range. The above results all combine to indicate that the distortion of the amplifier with feedback is held to an exceedingly low value over the entire working band of the amplifier.

In conclusion, a word of justification is desirable for building an amplifier with 65w of output power (or 130w peak, as the advertisements say!). The peak rms sound intensity level attained near an orchestra or chorus, a large organ, cymbals, or bass drum, is of the order of 110 db.[27] If this intensity level is to be reproduced in a 3,000-cubic-foot room, about 4.5 acoustic watts are required.[28] On assuming a speaker efficiency of 5 per cent, we see that the maximum rms peak power input to the speaker must be some 90w. Horn-loaded loudspeakers, which can be more efficient than 5 per cent would require less input power to produce the above level. Finally, it is rarely desirable to reproduce exactly the maximum original level unless the room is very large or its walls reinforced. These considerations indicate that an average output power of 20w to 30w would be quite sufficient for the vast majority of applications. Since the bass drum has its maximum acoustical output power below 60 cps and the cymbals theirs above 8kc, it is particularly important, however, for true reproduction that the power-handling capacity of the amplifier not fall off appreciably near the ends of the working band. Flat power response between 30 and 15,000 cps would certainly be adequate.

For the exceptional applications such as filling large auditoria or driving magnetic disk cutters, the present amplifier, with its vanishingly small distortion, is a

[24] Langford-Smith, *op. cit.*, p. 613.

[25] Since the peak rms power in the output is four times the average rms power (equal to that for either of the input signals alone), an average power much greater than 15w cannot be used with the present amplifier when the CCIF method is employed.

[26] Langford-Smith, *op cit.*, p. 612.

[27] H. Fletcher, "Hearing, the determining factor for high-fidelity transmission," Proc. IRE, vol. 30, pp. 266–277; June, 1942.

[28] Langford-Smith, *op. cit.*, p. 864.

possible solution either in its present form or in a modification which eliminated some of the frills, such as dynamic and static automatic balancing. Such elimination would increase the distortion, but since it is initially so low, the final value should still be quite acceptably low.

Some of the salient amplifier features are summarized below, and we see that they more than meet the original design goals.

AMPLIFIER SPECIFICATIONS

Frequency and Power Response

1w output: −1 db down at considerably below 5 cps and at 72 kcps.

65w output: −0.5 db down at 19.8 cps and at 22.4 kcps.

Nonlinear Distortion

Less than 0.2 per cent intermodulation distortion up to 60w; 0.1 per cent at 45 w; 1 per cent at 67w.

Noise Level

−106 db referred to 60w output.

Voltage Gain

31 db.

Rise Time

3 μs.

Square-Wave Response

9 per cent tilt at 20 cps. No parasitic oscillations.

Special Features

Automatic circuits ensure continuous push-pull signal balance and static balance of output tube cathode currents. The incorporation of 30 db of negative feedback over three stages and the output transformer is accomplished without peaks in the response curve. Provision is made for adjustable positive or negative output current feedback and for local positive voltage feedback.

Properties of Junction Transistors

R. J. KIRCHER†

This is the first of a group of three tutorial papers on transistors, with special emphasis on use at audio frequencies, prepared by the Bell Telephone Laboratories Staff at the request of the editorial committee of the Transactions on Audio. The other two articles, "Design Principles of Junction Transistor Audio Amplifiers," by R. L. Trent, and "Design Principles for Junction Transistor Audio Power Amplifiers," by D. R. Fewer, will appear in succeeding issues of this publication.—*The Editor.*

***Summary*—The motion of electrons and holes is considered in relation to the PN junction and it in turn is considered in relation to the junction transistor. Electrical properties, equivalent circuit diagrams, and limiting conditions of operation of junction transistors are discussed. Special equations and features of the common base, common emitter, and common collector configurations are developed and tabulated.**

INTRODUCTION

THE TRANSISTOR is the result of intensified research in the domain of solid state physics following World War II. From this work Shockley, Bardeen and Brattain conceived the idea that amplifying properties should be obtainable from semiconductors. This concept became a reality with the announcement of the point contact transistor in 1948.[1] The certainty that an amplifier could be made from a semiconductor element gave a tremendous impetus toward the development of an amplifying structure which would not require point contacts. By 1951 this objective was realized with the announcement of the junction transistor.[2] With the development of the junction transistor communications engineers, and particularly audio engineers, have received a new and versatile electronic device. The stature of the junction transistor grows on reviewing its unique features. Most striking is the property that electronic amplification occurs within a solid substance. This action is realized without the equivalent of heater or cathode power, so that it is instantaneous. Amplification comparable to that of a pentode electron tube is obtained for electrode

† Bell Telephone Labs., Murray Hill, N. J.

[1] J. Bardeen and W. H. Brattain, "The transistor, a semiconductor triode," *Phys. Rev.*, vol. 74, pp. 230–231; July 15, 1948.

[2] W. Shockley, M. Sparks, and G. K. Teal, "*P-N* junction transistors," *Phys. Rev.*, vol. 83, pp. 151–162; July 1, 1951.