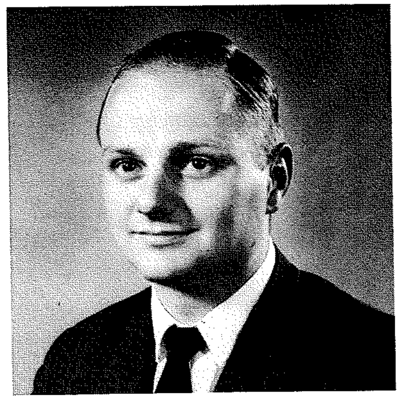
HIGH-QUALITY, LOW-COST, 15-WATT, COMPLEMENTARY-SYMMETRY POWER AMPLIFIER

This high-quality, low-cost audio power amplifier features a transformerless, direct-coupled, complementary-symmetry driver-output circuit. The class B output stage develops an RMS power output of 10 watts into an 8-ohm load (15 watts into 4 ohms), and has an EIA rating (stereo) of 30 watts into 8 ohms.



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THE primary objectives in the design of the 15-watt complementary-symmetry audio-amplifier system were to achieve a performance level superior to that previously attained, and at the same time to keep the cost at a level which would be competitive with present commercial designs. The success achieved in meeting these objectives is demonstrated by a review of the performance and cost of the system.

Performance of the 15-watt power amplifier is illustrated in Table I and Figs. 1 through 8. Typical harmonic-distortion levels are below 0.15% from 40 to 20,000 Hz both at full power output of 8 watts and at low power outputs. Intermodulation distortion is typically 0.1% at power levels of 8 watts or less. Other features include a hum and noise level of 94 dB below rated output, a frequency response (3 dB down) of 8 to 90,000 Hz, and a clean 20,000-Hz square wave with no ringing.

The low cost level of the amplifier results from three factors: 1) the system uses only 4 stages and no transformers other than the power transformer; 2) the first 3 stages use small-signal silicon transistors; and 3) the circuit is simple and straightforward and the number of electrolytic capacitors and diodes is very low.

ADVANTAGES OF TRANSFORMERLESS DESIGN

In order to achieve the desired cost and performance objectives, it was considered necessary to eliminate the driver transformer. A high-quality driver transformer is expensive and also has two major technical disadvantages. First, a rapid change in phase response to a total phase rotation of 180° at both high and low frequencies (when the transformer is capacitively coupled) makes it impossible to stabilize high amounts of feedback. (Feedback stability criteria are analyzed in Ref. 2, *RCA Application Note* No. AN-3185, which describes this amplifier.) Second, magnetic radiation of the power transformer causes a hum pickup in the driver transformer.

Use of the right type of transformerless circuitry also makes it possible to eliminate two major problems associated with capacitive coupling to the speaker. One problem is that the natural unbalance of the system prevents ripple cancellation at the speaker. The second is that the center voltage may go off-center under drive and cause premature clipping because there is no direct control over this voltage. Both problems are eliminated in the complementary-symmetry system when a high level of DC feedback is used to hold the center voltage at the proper point and a high level of AC feedback is used to cancel the ripple signal.

REQUIREMENTS OF COMPLEMENTARY SYMMETRY

Pure complementary symmetry is the only type of transformerless output circuitry that has very good thermal stability and thus permits the use of germanium output devices. (*Pure* complementary symmetry should not be confused with *quasi* complementary-symmetry circuits, which have poor thermal stability and thus can normally be used only with all silicon devices. These circuits also have relatively higher circuit complexity.) Because pure complementary symmetry also has a high performance level and low circuit complexity, it was selected as the best circuit configuration for this application.

The key to success of this complementary-symmetry design is the use of an *n-p-n* output transistor that complements the high-performance, high-frequency 2N2148 *p-n-p* power transistor. Three significant characteristics are necessary for the *n-p-n* complement: 1) a gain-bandwidth product f_T nearly the same as that of the 2N2148; 2) a minimum beta (measured at low collector-to-emitter voltage and peak collector current) high enough to maintain reasonable loop feedback and linear operation in the driver stage; and 3) static

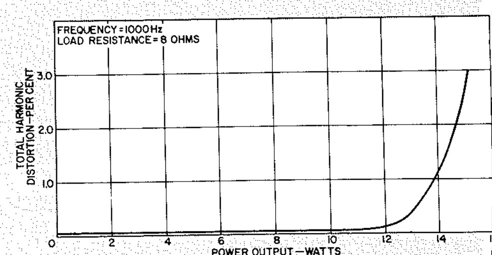

Fig. 1—Total harmonic distortion as a function of power output.

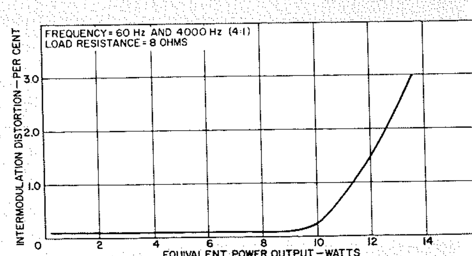

Fig. 2—Intermodulation distortion as a function of power output.

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characteristics at least as good as those of the 2N2148. A mismatch in beta or "fall-off" characteristics of the two output devices is not significant because of the high degree of feedback employed. Either a germanium or a silicon transistor can be used, therefore, provided it has the required characteristics of matched f_T, high minimum beta, and adequate static characteristics.

The RCA Dev. No. TA2577A used as a complement for the 2N2148 is a silicon power transistor in which the base region is grown epitaxially to form the collector-base junction. The emitter is then diffused into the base region to form the emitter-base junction. This type of construction results in the desired f_T of 3 to 5 MHz, a minimum beta of 50 at a collector-to-emitter voltage of 1 volt and a collector current of 2 amperes, and static characteristics equal to or better than those of the 2N2148.

Gain-Bandwidth Product

It is important that the two output transistors have matched f_T's for feedback-stability considerations. If the f_T's are not nearly the same, the phase-correction capacitor in the feedback cannot properly correct for both output units and ringing or oscillation will result on one half of the output.[2] A mismatch in f_T also causes very distorted output at the point where the gain of the lower-frequency transistor starts to decrease. The match in response of the 2N2148 and the TA2577A is illustrated by the 200-kHz sine-wave output shown in Fig. 5. The 20-kHz square wave shown in Fig. 6 demonstrates that no ringing occurs on either half of the square wave.

It is also desirable that the actual f_T's of the two devices be in the vicinity of 5 MHz. Feedback-stability rules[2] dictate that the response of one stage within a loop be inferior to that of the rest. Because the output transistors are usually the limiting factor in high-frequency response, the output-stage response should be as low as possible without materially affecting the 20-kHz distortion. An f_T of about 5 MHz provides the best results. In addition, higher-f_T devices tend to have poorer transient breakdown capability, and it would not be desirable to sacrifice transient-handling capability for unnecessary added high-frequency response. The TA2577A, for example, can handle pulses up to 100 watts (at collector-to-emitter voltages up to 25 volts) for periods as long as 50 milliseconds.

Beta

The minimum beta of 50 at peak current makes it possible to limit driver dissipation to 1.5 watts. Thus a metal-case, small-signal silicon transistor with a heat sink can be used in the driver stage. Use of a small-signal transistor is desirable both from a cost standpoint and because, as mentioned previously, the response in the driver stage should be very high compared to that of the output stage. The driver stage for the 15-watt complementary-symmetry amplifier must have a collector-to-emitter breakdown voltage rating $V_{(BR)CEO}$ of 40 volts and must be capable of linear operation at currents up to 200 milliamperes with a low saturation voltage $V_{CE(sat)}$. The 2N4074 *n-p-n* silicon epitaxial driver transistor is suitable for this type of operation.

CIRCUIT DESCRIPTION

Fig. 9 compares a block diagram for the complementary-symmetry power amplifier with that of a typical power amplifier using a driver transformer. These diagrams show that the front end (*Q1* and *Q2*) is essentially the same in either system. The major differences in the complementary-symmetry system are that the driver transformer is replaced by an additional transistor with a feedback loop and that the output uses a complementary pair of transistors rather than two transistors of the same polarity.

If the two systems shown in Fig. 9 were to have approximately equivalent cost levels, the cost of the output transistors would have to be the same for both systems, and the cost of the additional transistor *Q3* would have to match the cost of the driver transformer it replaces. If the cost of *Q3* is less than that of the driver transformer (the pentafilar-wound transformers used in high-quality systems are expensive items), the savings can be added to the cost of the complementary pair to obtain equivalent systems cost.

The complementary-symmetry system has improved technical performance, primarily because of the additional AC

Fig. 3—Total harmonic distortion as a function of frequency.

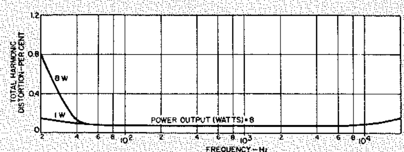


Fig. 4—Sine-wave output at a frequency of 5 Hz (vertical scale = 2V/cm, horizontal scale = 50 ms/cm).

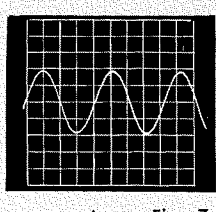


Fig. 5—Sine-wave output at a frequency of 200,000 Hz (vertical scale = 2V/cm, horizontal scale = 1 μs/cm).

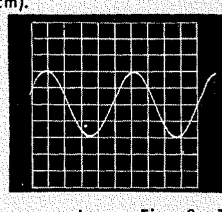


Fig. 6—Square-wave output at a frequency of 200,000 Hz (vertical scale = 2V/cm, horizontal scale = 10 μs/cm).

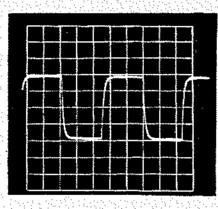


Fig. 7—Square-wave output at a frequency of 20 Hz (vertical scale = 2V/cm, horizontal scale = 10 ms/cm).

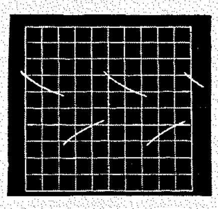


Fig. 8—Tone bursts at a frequency of 1000 Hz (vertical scale = 2V/cm, horizontal scale = 0.5 s/cm).

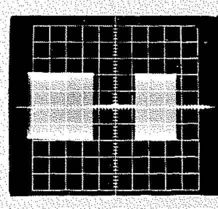


TABLE I—Performance of 15-Watt Complementary-Symmetry Power Amplifier

Power Output:	
At 1000 Hz for harmonic distortion = 5%	15 watts
At 1000 Hz for harmonic distortion < 0.15%	10 watts
From 40 to 20,000 Hz for harmonic distortion < 0.15%	8 watts
Distortion Levels:	
Total harmonic distortion as a function of power output at 1000 Hz	See Fig. 1
Intermodulation distortion as a function of power output at 60 and 4000 Hz (4:1)	See Fig. 2
Total harmonic distortion as a function of frequency at a power output of 8 watts	See Fig. 3
Frequency Response:	
At an output of 1 watt:	
1 dB down	20 to 50,000 Hz
3 dB down	8 to 90,000 Hz
Power Bandwidth:	
At an output of 8 watts for total harmonic distortion of 5%	15 to 100,000 Hz (See Figs. 4 and 5)
Sensitivity:	
For power output of 8 watts at 1000 Hz	160 mV into 3300 ohms
Hum and Noise, input open	94 dB below 8 watts
(Note: proper grounding and shielding of input circuit is necessary to obtain this measurement.)	
Electrical Stability:	
20,000-Hz square wave	See Fig. 6
20-Hz square wave	See Fig. 7
1000-Hz tone bursts	See Fig. 8

All measurements made at AC line voltage of 120 volts with 8-ohm load and one channel operating, using internal power supply shown in Fig. 10. Performance with 4-ohm or 16-ohm load is essentially the same except that power output at the clipping level is increased slighty with 4 ohms and decreased slightly with 16 ohms.

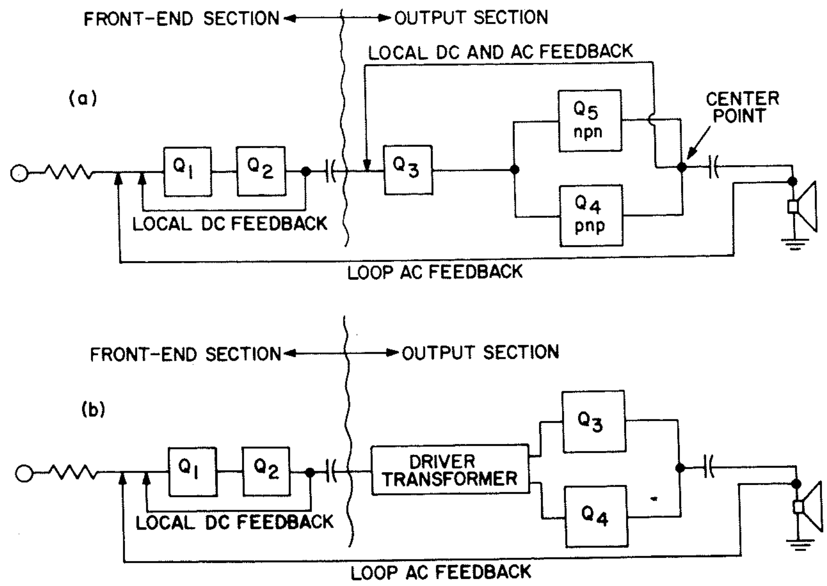


Fig. 9—Block diagrams of (a) a complementary-symmetry system, and (b) a driver-transformer system.

and DC feedback loop from the center point of the output stage to the base of the driver *Q3*. This feedback provides the following technical advantages: 1) the DC feedback from the center point holds the center voltage at the desired bias point and prevents premature clipping as a result of variations in output transistors; 2) the ripple level (which might otherwise be a problem in an unbalanced, single-supply output stage) is greatly reduced; 3) distortion levels are greatly reduced (so that linearity and matching of the output devices are not very significant); 4) frequency response is greatly improved; and 5) high-frequency phase shift can be held close to 90° (if the frequency response of *Q3* is very high) at frequencies up to the point where open-loop gain is less than unity (it is then much easier to stabilize the loop feedback). The additional feedback does not result in a lower overall current gain (from the output of *Q2* to the load) because the transistor *Q3* has a much higher current gain than the transformer it replaces.

CIRCUIT OPERATION

A schematic diagram of the complementary-symmetry power amplifier is shown in Fig. 10. Operating fundamentals for the complementary-symmetry driver-output stage are described in the RCA Application Note[2] mentioned earlier. The idling current in the output stage is established by the voltage drop across the 1N3754 silicon bias diode *D1*, and is stabilized by the two 0.51-ohm emitter resistors *R11* and *R12*. The DC drop across the bias diode is virtually independent of changes in the current through it (i.e., the diode has a low dynamic impedance). However, this voltage decreases with increases in temperature, and thus partially compensates for changes in the base-to-emitter voltage of the output transistors. As a result, the idling current is extremely stable. With the output transistors used, the single bias diode provides an output idling current of about 10 to 20 milliamperes. This low idling current does not create a crossover distortion problem because the output stage is driven from a high AC impedance. The result is cool and stable operation in the output stage.

The idling current in the driver stage (which must at least equal the maximum peak base current required by the *n-p-n* output transistor) is established by the two 120-ohm bias resistors *R9* and *R10*. The driver current is equal to the difference between the supply voltage and the center voltage divided by the series resistance ($R9 + R10$). and is about 92 milliamperes.

For proper operation of the circuit, the current shown as I_1 in Fig. 10 must remain essentially constant during AC excursions of output voltage. For this purpose, a 250-microfarad "bootstrap" capacitor *C5* is connected between the bias resistors and the emitter of the *n-p-n* output transistor *Q5*. Because the voltage across the capacitor does not change during AC output-voltage excursions, the change in voltage at point *B* is the same as the change in voltage at point *A*. The change in voltage at point *C* is almost the same as that at point *A*, and differs only by the small change in the base-to-emitter voltage of *Q5*. Therefore, the voltages at points *B* and *C* change by essentially the same amount, the voltage across the 120-ohm resistor *R10* remains constant, and a constant current I_1 results.

The "bootstrap" capacitor *C5* is returned to point *A* rather than to the center point (as is the usual practice) to keep the change in voltage across the 0.51-ohm emitter resistor *R11* from appearing across resistor *R10*. When the change in voltage across *R11* appears across *R10*, the current I_1 is less constant and the dynamic-range requirements of the driver transistor *Q3* are increased.

It is important to note that a reverse base-to-emitter bias voltage is applied to the output transistors during the *off* half-cycles.[2] This condition also occurs in an output stage that uses a driver transformer, but not in many transformerless circuits. When reverse bias is applied to the output transistors during the *off* half-cycle, they are turned off in a very short time (the stored base-emitter charge is removed rapidly). As a result, the high-frequency operation remains in the highly efficient class B mode. When the output transistors are not reverse-biased, but instead are allowed to drift off, the stored charge holds the transistors on during part or all of the *off* half-cycle. Operation then shifts into class A push-pull, and efficiency is low and current drain, dissipation, and operating temperatures are high.

Bias for the base of the driver stage is derived from the center point of the output stage. As a result, a DC and AC feedback proportional to the center voltage is fed to the base of the driver stage. The actual DC center voltage which the feedback establishes depends on the ratio of resistors *R6* and *R7* and on the base-to-emitter voltage V_{be} and base current I_B of *Q3*. If a heavy direct current is bled through *R7*, changes in the base current of *Q3* become insignificant The DC voltage at the center point is then determined by the base-to-emitter voltage of *Q3*. Because the percentage of variation in V_{be} of a silicon transistor is small, the center-point DC voltage is held close to the desired value. The resistor values *R6* and *R7* must be chosen so that: 1) the bleeder current in *R7* is large compared to the base current in *Q3*;2) the ratio of the resistors provides the desired center-point voltage; and 3) the desired amount of AC feedback current is obtained.

ON-OFF TRANSIENTS

An important consideration in any type of output circuit, and especially in a transformerless circuit, is turn-on and turn-off transients. These transients must neither sound too objectionable nor cause operating conditions which would result in transistor failure.

In the circuit of Fig. 10, the speaker coupling capacitor *C6* charges through *Q5,* which is biased on, and the capacitor *C5.* As a result, there is a slight noise in the speaker, and the center-point voltage rises to its steady-state value in an underdamped manner. The turn-on transient is thus of a relatively short duration, and the operating conditions during the transient are well within the capabilities of the output devices.

After turn-off, *C6* discharges through *Q4,* which is also biased on. The resulting noise in the speaker is usually not noticeable because the amplifier continues to deliver output signal while the capacitor discharges. Again, no operating conditions occur which could cause transistor failure.

In some transformerless output circuits, including some variations of the basic complementary-symmetry circuit, voltage is obtained from a decoupled point in the power supply, rather than from the *B+* point right at the rectifier diodes. Because such a decoupled point comes up to voltage much more slowly than the point right at the diodes, the very long turn-on transient in this type of circuit can cause adverse operating conditions which can exist for as long as several seconds. As a result, one of the output transistors is biased to a high current for a period of time long enough to be essentially DC, and thermal runaway may occur during the turn-on transient at high ambient temperatures.

FEEDBACK LOOP

The front end of the power amplifier shown in Fig. 10 consists of a pair of *n-p-n* silicon transistors in a direct-coupled circuit. The feedback from the emitter of *Q2* to the base of *Q1* serves primarily to hold the DC operating point of *Q2* within the limits necessary to prevent a dynamic-range limitation, despite variations in individual transistors or in temperature. The loop feedback resistor *R8* also serves as the DC bias resistor from the base of *Q1* to ground (the resistor returns to ground through the output voltage-divider resistors *R13* and *R14*).

For high-frequency stabilization of loop feedback, it is necessary that one element within the loop have relatively poorer high-frequency response than the rest.[2] Because of the local feedback established by resistor *R6,* the response of the output section is very high. Therefore, the response must be limited somewhere in the front-end section. The collector-to-base feedback capacitance C_{ob} of transistor *Q1* can be used for this purpose. The high collector load impedance of *Q1* causes a high-frequency local feedback current to flow through this capacitance and thus limits the high-frequency response of *Q1.*

The same rules hold for low-frequency loop-feedback stabilization. It is necessary that one element within the loop have relatively poorer low-frequency response than the rest. The low-frequency response is limited by *C6,* the speaker coupling capacitor.

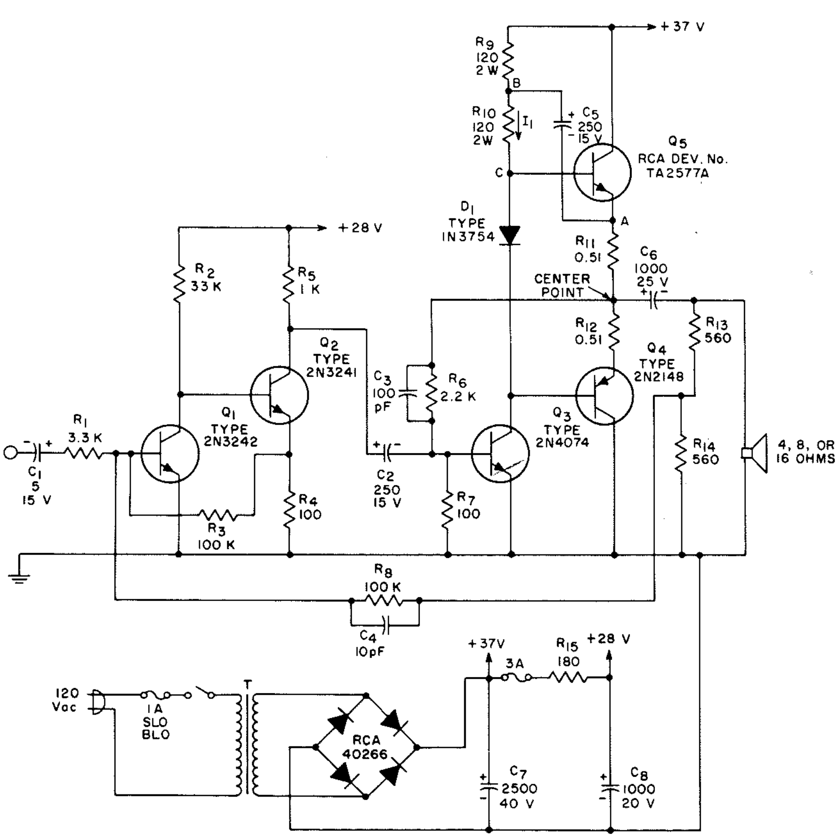
Because the value of resistor *R8* is established by the DC bias considerations for the front end, the proper amount of AC loop feedback is established by deriving the feedback from a voltage divider across the output. Resistors *R13* and *R14* divide the output voltage down to a level which provides the desired feedback current through *R8.* Resistors *R13* and *R14* also serve as an output termination when there is no speaker load.

It is important to note that, because there is no driver transformer, negative feedback can be applied only to certain points in the circuit. (When a driver transformer is used, the primary can be phased either way to provide the proper phase to the feedback signal, wherever it is located.) The points that can be used in the circuit of Fig. 10 are the base of the driver transistor *Q3* (which is used for local feedback), the emitter of *Q2,* and the base of *Q1* (which is used for the loop feedback).

When feedback is applied to the base of a transistor, the transistor must be driven by a current source. If a resistor is used to provide a current source, the resistor also isolates the feedback from the rest of the circuit. The 3,300-ohm resistor *R1* serves as a current source for *Q1* and prevents the feedback from interacting with the preamplifier circuit.

The high degree of feedback (about 32-dB local in *R6* and 34-dB loop in *R8,* for a total of 66 dB) results in an extremely low output impedance and a high degree of speaker damping. This large amount of feedback is the main reason for the extremely low hum and distortion levels in the amplifier. In spite of the large amount of feedback, the stability is excellent, as shown by the square-wave and tone-burst responses in Figs. 6, 7, and 8. This stability results from elimination of the driver transformer and careful observation of the rules of feedback stability (discussed, as noted before, in Ref. 2).

Fig 10—Schematic diagram of the 10-watt complementary-symmetry audio power amplifier.



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2. M. S. Fisher, *A High-Quality, Low-Cost, 15-Watt, Complementary-Symmetry Power Amplifier,* RCA Application Note AN-3185, June 1966. (Available from Commercial Engineering. RCA Elec. Comp. & Devices, Harrison, N.J.)