A High-Power Amplifier With Minimum Distortion

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In an amplifier not for construction by the novice, the author tries for a new "ultimate" in performance.

THERE IS NO DOUBT that the best of present-day amplifiers conform to very high standards. However, a study of the problem of distortion indicates that still higher standards can be set for a distortion-free amplifier and the writer has evolved a design whose principal constructional feature is a new cross-coupled phase-splitter-driver.

In working on the new design the writer realized that a somewhat new approach was necessary. To begin with, the new circuit would have to satisfy these requirements:

(1) Negligible distortion — intermodulation, harmonic, phase and frequency.

(2) No tendencies toward low-frequency or ultrasonic oscillations.

(3) A maximum power rating far in excess of that required for the reproduction of the highest instantaneous peaks found in music.

(4) A negligible incremental output impedance, and a means for making this variable in the negative direction.

(5) Complete freedom from hum, hiss, and microphonics.

For the lowest distortion, any amplifier must be critically adjusted. Therefore, for the best results, it is recommended that the construction of the amplifier to be described should not be attempted by those who do not have access to at least an intermodulation analyzer, a vacuum-tube voltmeter and a good oscilloscope.

The Circuit

The basic philosophy inherent in the design was to make the amplifier as perfect as possible before the application of negative feedback, so that with its application the distortion would virtually vanish. The signal level inside the amplifier was to be kept at the lowest possible level to keep the distortion at a minimum. In order to realize this only one voltage feedback loop is used (from the output transformer secondary to the input cathode). Each stage must operate at as low a level as possible, which cannot be achieved easily with internal feedback loops. A minimum number of stages is used, and the circuitry kept simple. Incidentally, no matched components need be used.

The input stage (*Fig.* 1) is a triode whose circuit parameters have been experimentally adjusted to produce the lowest amount of intermodulation without feedback. This drives a modified cross-coupled phase-splitter-driver V_2-V_3-V_4. This new splitter-driver circuit is primarily responsible for the performance of the amplifier.

In the original cross-coupled phase splitter[1] there were two cathode followers driving two triode voltage amplifiers. For single-ended inputs the grid of one of the cathode followers was grounded and left unused. It was found that the unused cathode follower could be re-

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[1] J. N. Van Scoyoc, "A cross-coupled input and phase inverter circuit," *Radio Electronic Engineering Edition, Radio & Television News*, Nov. 1948.

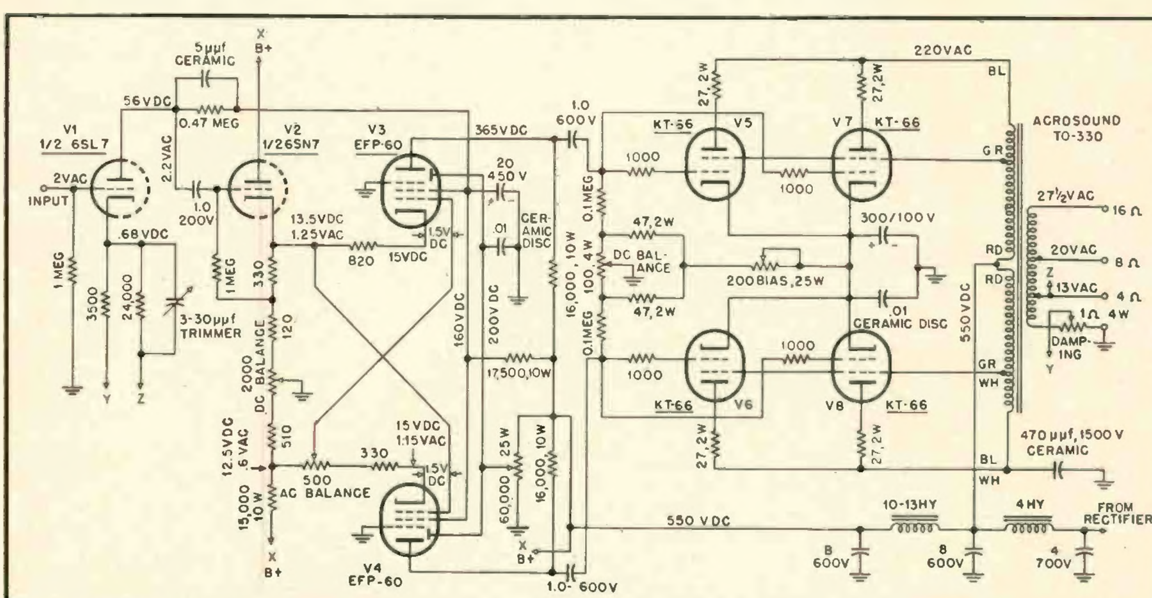

Fig. 1. Circuit diagram of the author's amplifier. Intermodulation and harmonic distortion are at lower than inaudible levels at 50 watts output. The EFP-60's are new, use a special 9-pin socket.

Fig. 2. Amplifier and power supply are together on the 13 × 17 × 3-inch aluminum chassis. Steel might be a better chassis material if the unit is to be handled to any extent.

placed with a resistor equal to twice the tube's nominal plate resistance with no effect at all upon the operation of the circuit except a slight additional amount of degeneration. This further helps to keep the circuit insensitive to the effects of drift and tube aging. This unused triode might then be used for the input voltage amplifier. However, tubes which make good low-distortion cathode followers in the cross-coupled circuit do not make good high-gain, low-distortion voltage amplifiers. Therefore, two separate tubes were used: one-half of a 6SL7 for the voltage amplifier and one-half of a 6SN7 for the cathode follower. The other halves of these tubes are unused and are thus free for other purposes such as a preamplifier.

Contrary to popular belief, a cross-coupled circuit is not self balancing, and the so-called balance controls incorporated in these circuits accomplish nothing but to change the d.c. biases on the voltage-amplifier tubes. Alternating-current balance *can* be accomplished by adjustment of this control to extreme values, thereby changing the mu of the tubes. This practice naturally leads to increased distortion. It must be realized that the drives to the power tubes are not necessarily equal to each other at the point of lowest distortion. Final tube unbalance and transformer unbalance, negligible in high-quality output transformers, account for this phenomenon. When final-tube and transformer unbalance cancel out the inverter unbalance, low distortion results, but this is not always the case.

Unbalance in the cross-coupled circuit can be traced to the driving of one tube by its cathode and the driving of the other by its grid. Assuming for the moment the absence of negative feedback (about 21 db of current feedback in the writer's circuit) the cross-coupled splitter reduces to a grounded-grid and a grounded-cathode amplifier driven by a cathode follower. In a grounded-grid stage the gain is equal to $\mu + 1$. In the grounded-cathode stage the gain is just μ.[2,3] When the gains of these two tubes are high they are approximately equal. There is one difference, however. Since the source is in series with the load in the grounded-grid stage, the tube has to deliver power. On the other hand, with the grounded-cathode stage, the input driving impedance is almost infinite for low frequencies. The cathode follower which drives these two stages has a low output impedance, but in order to get the least amount of degeneration from the cathode follower this output impedance cannot be reduced low enough to match the input impedance of the grounded-grid stage without drastically reducing the gain of the whole splitter.[4] Therefore, in order to get any gain out of the circuit there has to be a mismatch between the cathode follower and the grounded-grid stage, with consequent unbalance of drive to the two stages.

To adjust the drives to the final tubes for lowest over-all distortion a method had to be devised to vary the drives to the grounded-grid and grounded-cathode stages of the splitter. To see how this was accomplished refer to *Fig.* 1 and note the 500-ohm potentiometer in series with the cathode-bias resistor of V_4. The arm of this potentiometer is connected to the grid of V_3. Part of the negative feedback applied to V_3 is developed across the cathode resistor of V_4. By varying the amount of degeneration applied to the V_3 grid (by tapping down on the V_4 cathode resistor) we can equalize the loss of drive to the upper tube due to its mismatch. With this control, the a.c. balance of the inverter can be varied about 10 per cent either way—enough to take care of any final-tube or transformer unbalance. Adjustment of this control affects the d.c. balance of the inverter slightly, necessitating readjustment of the 2000-ohm d.c.-balance potentiometer.

In order to comply with the original requirements of a minimum number of stages and of keeping the a.c. voltages inside the amplifier small, a new tube just recently placed on the market, the EFP-60* was used as the voltage-amplifier-driver in the cross-coupled stage. This tube utilizes the principle of secondary emission to obtain the extremely high transconductance of 25,000 micromhos. Using this tube, one can use a very low value of plate-load resistor, thereby keeping the driving impedance low, and still obtain a very high gain. In the splitter-driver circuit, the tubes have a gain of better than 250 with a 16,000-ohm plate load resistor. Their circuit parameters have also been experimentally adjusted for lowest distortion. Naturally, using such high-gain tubes presents some problems of input-output isolation and shielding, but by careful construction one will have no trouble if the usual precautions adopted when working with high-gain, wideband amplifiers are observed.

A 60,000-ohm variable resistor across the high-voltage supply furnishes the "negative current" (a term adopted by the manufacturer) for the secondary cathode of the EFP-60. One electron impinging on this secondary cathode knocks out 5 more. These electrons are supplied from ground, while the electrode is kept at about 200 volts above ground by the B+ supply. This means that the resistor slider is kept close to its grounded end. A ceramic disc capacitor bypasses the secondary cathodes to ground at the sockets of the EFP-60's; this helps to stabilize the stage. In this circuit the suppressors of the EFP-60's are connected to ground instead of to their cathodes. It was found that when they were connected to the cathodes distortion and parasitic oscillations resulted. The fact that they are about 15 volts negative with respect to the cathodes has no effect on the operation of the tubes.

The Power Stage

The output stage is a conventional push-pull-parallel one with KT-66's ultralinear-connected, using the Acrosound TO-330 transformer. A 300-μμf capacitor and shunt .01-μf ceramic disc bypass the final cathodes to ground. It is essential to use this much capacitance to keep high-level intermodulation at a minimum. The disc capacitor nullifies the effects of inductance in the large electrolytics, thereby providing adequate bypassing over the entire useful frequency spectrum.

A 470-μμf high-voltage ceramic capacitor from the blue-white plate lead of the output transformer to ground corrects for a capacitive unbalance to ground which would otherwise exist in the transformer due to the geometry of the windings. This capacitor clears up a slight ultrasonic parasitic which developed when the transformer was overloaded. The measured capacitive unbalance existing in the transformer is actually less than 100 μμf. The 470-μμf capacitor evidently has some other stabilizing effect than just cleaning up the capacitive unbalance to ground.

[2] F. E. Terman, "Radio Engineers Handbook," 1st ed., McGraw-Hill, New York, 1943, pp. 470–471.

[3] "Reference Data for Radio Engineers," 3rd ed., Federal Telephone and Telegraph Corporation, 1949, pp. 252–253.

[4] S. W. Amos "Valves with resistive loads," *Wireless Engineer*, April, 1949, pp. 119–123.

* The EFP-60 tubes are manufactured in the Netherlands by N. V. Philips Gloeilampenfabrieken, Eindhoven, and are imported into the U.S.A. by Amperex. Usually stocked by the larger parts houses, they may be ordered from Amperex, Hicksville, N. Y. if unavailable in your locality. The socket is Amperex No. S-13211.

The negative feedback voltage is taken from the 4-ohm tap on the transformer secondary. It was found experimentally that in this way more feedback could be applied without instability (high-frequency oscillation) than could be applied when the feedback was taken from the 16-ohm tap, the conventional practice. The negative feedback is applied to the cathode of the 6SL7 voltage amplifier V_1 through a phase-advance network consisting of a 3-30-μμf trimmer capacitor in shunt with a 24,000-ohm resistor. The amplifier has a voltage gain of about 535 from the input to the grids of the KT-66's with no feedback. 32 db of negative feedback is applied in one loop from the output to the input. With this amount of degeneration, 2 volts at the input drives the amplifier to 50 watts output, so that there is a voltage gain of about 10 to the 8-ohm tap on the output transformer secondary.

Since the feedback loop includes the output transformer, it is natural to expect stability problems to arise. The amplifier is perfectly stable at the low-frequency end, due to the large ratio (greater than 400 to 1) of the coupling capacitor low frequency cutoffs.[5] There is however another low-frequency cutoff, that of the transformer. Its effect on the low-frequency stability is hard to ascertain because it varies with the level of transformer excitation. Suffice to say no low-frequency peak was found in the range ending at 1 cps. Due to the difficulty of making measurements, the region below 1 cps was not investigated. The long time constants used in the amplifier reduce the low-frequency phase shift to almost zero, while extending the low-frequency response of the amplifier to below 1 cps. This extended response is obtained without the low-frequency transient instability peculiar to Williamson-type amplifiers. Many of these amplifiers are in a condition of near oscillation at a frequency of about 2 cps and when the amplifier is excited by a low-frequency peak, it is "triggered" at this subsonic frequency, causing the reproduced sounds to have a muddy, mushy, lack-of-definition quality. These difficulties are completely absent with the amplifier discussed here.

[5] N. H. Crowhurst "Feedback," "Audio Handbook No. 2," Norman Price, Ltd., 1952.

At extreme high frequencies, the large amount of feedback employed tends to introduce problems. Since more than the 30-db limit recommended for the transformer is being applied, it is natural to expect some difficulties to develop. In this amplifier two unstable regions were found—regions where the phase lag of the amplifier was sufficient to cause it to oscillate through the feedback loop. These were 110 kc and about 1.5 mc.

The instability at 1.5 mc. was eliminated by reducing the loop gain at this frequency to a value less than unity, shunting the 470,000-ohm plate load resistor of V_1 with a 5-μμf ceramic capacitor. This expedient did not work, however, for the 110 kc oscillation. It was found that the output transformer leakage reactance was causing sufficient phase delay at this frequency to prevent 32 db of stable feedback. Various "step" selective phase-advance networks were tried in order to steer the phase characteristic of the amplifier away from the 180-deg. point in this region. None of these "step" networks had much effect, so a variable 3-30-μμf mica trimmer across the feedback resistor was finally adopted. This expedient cured the oscillation.

An adjustable incremental damping factor network has been incorporated into the design of the amplifier. The operation of this network is to control the amount of negative feedback applied to the amplifier by means of the current flowing in the load. The adjustment of the 1-ohm potentiometer is conventional.[6] If not desired, this network can be eliminated without affecting the performance of the amplifier in the slightest.

Construction

The amplifier and power supply, illustrated in *Figs*. 2 and 3, were constructed on the same 13 × 17 × 3-in. aluminum chassis. It is recommended that the amplifier be laid out and constructed in straight-line fashion. Particular attention should be paid to common stage grounds and bypassing the EFP-60 dynodes to ground directly at the socket. Radio-frequency wiring practices are advisable. It is recommended that the filament leads be shielded or twisted and kept close to the chassis. All signal wiring should be point-to-point and kept away from the chassis. In particular, the 1-μf coupling capacitors should be suspended away from the chassis by their pigtail leads. It is definitely *not* recommended that bathtub capacitors be used for this purpose. The writer suggests metallized-paper tubular capacitors for their small size, high voltage rating and self-healing properties. Only non-inductive composition resistors should be used for the grid and plate stoppers of the KT-66's, and these should be connected by the shortest leads directly to the sockets. The KT-66's should be allowed plenty of ventilation, as they must be run at maximum ratings for lowest distortion. The filament supply for the amplifier must be the correct 6.3 volts; the EFP-60's are very sensitive to deviations in filament supply voltage, especially in the negative direction. The plate supply should be able to deliver a continuous 300 ma at 550 volts with good regulation. A pair of 5V4's as rectifiers is recommended. Although the manufacturer's recommended maximum plate voltage per plate is exceeded, no trouble has been experienced in several months of extended operation. It was found that the better regulation inherent with 5V4's will reduce the high-output-level intermodulation by a factor of more than 25 per cent over a pair of 5R4GY's.

[6] "A new approach to loudspeaker damping," Warner Clements, AUDIO ENGINEERING, August, 1951.

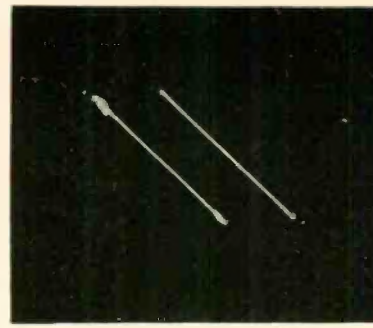

Fig. 4. Fuzzy tips on the input-output characteristic shown in this oscillogram denote instability at extreme high frequencies at very high power levels. Adjust the feedback trimmer to eliminate this trouble.

Adjustment

Perfection is no accident; distortion-free operation cannot be achieved without critical adjustment.

After checking the wiring the following adjustments are made with the power off: Set the d.c. balance control to its middle position, the arm of the a.c. balance potentiometer close to the junction of the 15,000-ohm and 510-ohm resistors, and the slider on the 60,000-ohm resistor to the grounded position. Adjust

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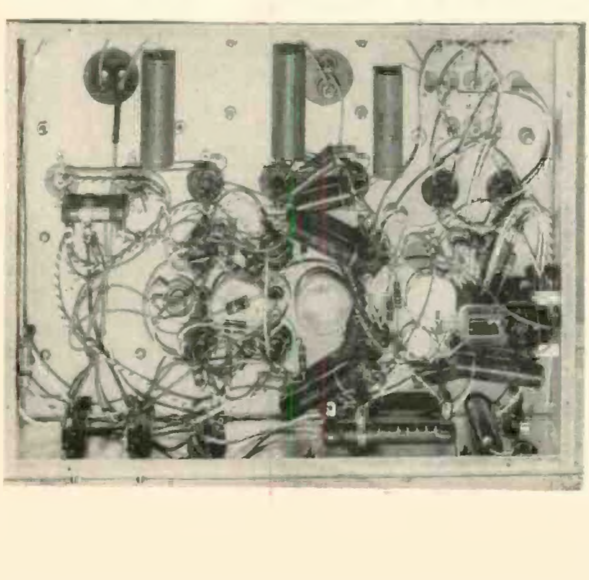

Fig. 3. Under the chassis of the amplifier. Despite experimental nature of this model the important leads are dressed as the author recommends in the text.

the 100-ohm final-tube plate-current balance control to its middle position, set the slider on the 200-ohm final-stage bias resistor nearest the cathode end and screw up fairly tightly the trimmer across the 24,000-ohm feedback resistor.

Apply plate and filament power. After 5 or 10 minutes warmup, adjust the dynode voltage for the EFP-60's. With a d.c. voltmeter connected from dynode to ground, slowly move the slider on the 60,000-ohm resistor away from the grounded end. As the slider is moved, the voltage will rise slowly at first, and then jump abruptly to a value of about 170 volts. Leave the slider in this position for the time being. The sudden jump in voltage means that the tubes have commenced their secondary emission and the secondary cathodes are drawing current.

Adjust the d.c. balance by placing the voltmeter across the plates of the EFP-60's and adjusting the 2,000-ohm d.c. balance control until there is no voltage differential. There will be some constant random fluctuations of the meter needle due to supply and tube fluctuations. These should be disregarded.

Connect the intermodulation analyzer and adjust the amplifier for about 50 watts into a noninductive load. Adjust the 500-ohm a.c. balance control for lowest intermodulation at this power level. Remove the KT-66's from their sockets. Feed a very small 60-cps signal to the input, and adjust the signal's magnitude until a signal of about 25 volts is measured at the output of the EFP-60's with an a.c. v.t.v.m. Be careful to isolate the test leads from the amplifier input or feedback oscillations will result. Readjust the dynode voltage of the EFP-60's for maximum gain, i.e., maximum voltage output. Leave the slider set at that position. Readjust the d.c. balance. Replace the KT-66's *in the same sockets as before*, reconnect the intermodulation analyzer, and again adjust the a.c. balance for lowest intermodulation. Balance the static d.c. plate currents in the output transformer by adjusting the 100-ohm final balance potentiometer for zero voltage differential across the plates of the KT-66's. This adjustment is very important as d.c. magnetization of the output transformer core causes its low-frequency characteristics to deteriorate badly.

Connect an oscilloscope to display the input-output characteristic of the amplifier at 1000 cps. Increase the input to the amplifier until it is on the verge of overload. Adjust the trimmer across the feedback resistor to eliminate the parasitic oscillations at the tips of the linearity characteristic (*Fig.* 4). There should be a fairly broad range to this adjustment. Reconnect the intermodulation analyzer. Adjust the input for 50 watts output and finally readjust the cathode bias on the KT-66's (with the modulation. These adjustments, if carried out properly, should yield an intermodulation distortion figure of less than .09 per cent at 50 watts, using 60 and 7000 cycles mixed 4:1. It should be noted that the a.c. balance is a fairly critical adjustment for lowest distortion and should be readjusted periodically as the tubes age. Whenever the a.c. balance is adjusted, the d.c. balance should be touched up. It is important that during all these adjustments the 1-ohm damping-factor control should be set with its moveable arm grounded.

Operation

The writer has found that with different speaker systems the adjustment of the 3-30-μμf trimmer across the feedback resistor might have to be changed to insure the greatest stability at high power levels.

It is advisable that a high-pass filter be incorporated somewhere preceding the amplifier. The extended low-frequency response of this amplifier, coupled with the fact that *the output transformer power handling capacity drops off at the rate of twelve db per octave below twenty cps,* makes the amplifier especially susceptible to overload at frequencies below, say, 10 cps. It is conceivable that record eccentricities, rumble, etc., will cause the output transformer to saturate at these extremely low frequencies, thereby generating distortion and overloading the loudspeakers.

It was found that at high power levels, parasitic oscillations appearing on the tips of the linearity characteristic caused the intermodulation to rise to very high values. It seems that even though the frequency of these oscillations is way above the audio spectrum, they have a very drastic effect on the distortion characteristics of the amplifier, and their complete elimination is of paramount importance. Even faint wisps and wiggles seen only on a wideband, triggered-sweep oscilloscope have a great effect on the distortion. The mechanism of these effects has yet to be investigated.

The harmonic distortion of the amplifier is less than 0.5 per cent at 20 cps at 50 watts. It decreases rapidly as the frequency is raised, and is too small to be accurately measured at 100 cps at 50 watts. At no frequency between 100 cps and 20 kc does the harmonic distortion exceed .05 per cent at the 50-watt level.

Some experimentation has been carried out with the use of tubes other than KT-66's. 807's, 5881's, and the new Tung-Sol 6550's have been tried. Of these, the 6550's are the most promising. The 5881's cannot deliver the high power without being overloaded, and the 807's, although capable of delivering more power than the KT-66's, are very susceptible to r.f. parasitics and are not recommended. A single pair of 6550's can deliver almost as much power as four KT-66's, and they seem to be much freer from parasitics than even the KT-66's. Since the Acrosound TO-330 is not matched to these tubes, maximum power, in this circuit, is delivered only when a different value of load resistance from the nominal values of 4, 8, and 16 ohms is used. It was found that with a load of about 11 ohms at the 16-ohm tap, the amplifier delivered the most power—in this case about 68 watts maximum. The distortion with two 6550's was not quite so low as was obtained with four KT-66's.

The author is greatly indebted to Mr. David Hafler of Acro Products Company for his many suggestions, and in particular for the suggestion to eliminate one of the cathode followers in the cross-coupled inverter. Also the author would like to thank Professor Earl W. Keller of the M.I.T. Electrical Engineering Department for the generous donation of his time, advice, and encouragement.