Zero-Impedance Output Stage

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Excellent transient and low-frequency response and good loudspeaker damping make this amplifier suitable for high-quality, low-power applications.

THE ZERO-IMPEDANCE STAGE to be described was designed for home use, along with its driver, to give good quality performance at moderate cost. A series R-C circuit (R_1, C_1 in *Fig.* 1) shunts the primary of the output transformer so that the output tube works into almost unity power factor load. This minimizes harmonic distortion and phase shift. The feedback circuits are direct coupled and the negative voltage feedback is taken from the primary of the output transformer rather than the secondary in order to reduce undesired phase shift to a minimum in this feedback loop.

The low-frequency response is exceptionally good because the stage is effectively acting as a zero-impedance source feeding the primary of the output transformer. The output transformer used was of good quality and had 1-inch stack. A frequency response taken with the loudspeaker connected, and measuring output voltage across the secondary of the output transformer indicated the 3-db-down point was below 20 cps at the low end, and at 5000 cps at the high end, and only 9 db down at 15,000 cycles per second. At 2½ watts output into a resistance load at 400 cps, the total r.m.s. distortion was under 5 per cent. This is relatively high by most standards, but quite low for a 6V6.

A disadvantage of taking the negative voltage feedback from the primary of the output transformer is that this feedback cannot correct for the fall-off in high-frequency response in the transformer. The writer prefers to sacrifice some high-frequency response for minimum phase shift in the negative feedback circuit. This assures that the feedback works most effectively, reducing intermodulation distortion to a minimum, giving maximum reduction of harmonic distortion and maintaining a low-impedance source feeding the output transformer, over and beyond the complete audio frequency spectrum. It is possible to compensate for this loss in highs by a fixed equalizer in the preamplifier, but this was not done because the high-frequency loss was not serious. Most preamplifiers incorporate some form of tone control circuit with treble boost which

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can be used for this equalization.

The use of the series R-C network across the primary of the output transformer to provide virtually unity power factor load to the tube is not new. The theory of this is well known. If a series *R-C* circuit and a series *R-L* circuit are connected in parallel as shown in *Fig.* 2, where the *R*'s are equal, it can be proved that the impedance of the parallel combination will be a pure resistance equal to R at all frequencies, if $R = \sqrt{L/C}$. At frequencies above the resonant frequency of the loudspeaker, the impedance measured across the primary of the output transformer with the loudspeaker load on the secondary may be roughly approximated by a series *R-L* circuit. Consequently, within this frequency range, which extends from approximately 125 cps to the highest audio frequencies, the composite load impedance presented to the tube is very nearly pure resistance with small variation in magnitude with frequency.

If values of R_1 and C_1 are chosen to give the optimum composite load impedance, there will be appreciable reduction in available output power at the higher frequencies where the reactance of C_1 becomes small in comparison with the magnitude of R_1. This is a serious disadvantage. A compromise between these two factors was made in this design.

Adjustment of R and C

The effect of changing R_1 and C_1 can be observed readily on an oscilloscope by the simple circuit of *Fig.* 3, and the values of R_1 and C_1 were finally selected in this way. The value of R_p used was 47,000 ohms, which approximates the plate resistance of the 6V6. The phase angle of the combination is determined from the ellipse appearing on the screen.

For the tube operating voltages used, the load impedance Z presented to the tube should be from 7000 to 10,000 ohms. The output transformer should match the loudspeaker to this load. A good compromise for R_1 is 2 to 4 times the magnitude of Z and the time constant R_1C_1 should be approximately 150 microseconds.

The value for Z used by the author was low, being around 5,000 ohms; R_1 was 15,000 ohms and C_1 was 0.01 µf. These values are not critical.

For R_1C_1 time constant of 150 microseconds, the reactance of C_1 is equal to R_1 at approximately 1000 cps. Consequently the power loss in the resistor is low below this frequency. In music, very few fundamental tones occur above this frequency which corresponds to two octaves above middle "C" on the piano. Even though some available power is lost at higher frequencies, it is not a serious matter.

In the usual 6V6 class "A" power amplifier, the average screen current and the average plate current both increase when a signal of sinusoidal or symmetrical waveform is applied. If the signal is keyed on and off and a d.c. milliammeter placed in the plate circuit, the meter fluctuates wildly. If instead of maintaining a constant screen voltage, a particular value of screen dropping resistor R_2 is chosen, it is possible to minimize this fluctuation in d.c. plate current provided the screen bypass capacitor is removed. The action is simple. In the presence of signal, screen current increases. The resulting increased voltage drop in R_2 causes a reduction in screen voltage just sufficient to provide the required compensation. In order to obtain instant action, the screen-bypass capacitor must be removed. The removal of this capacitor results in about 10 per cent reduction in voltage gain, and a loss in screen filtering.

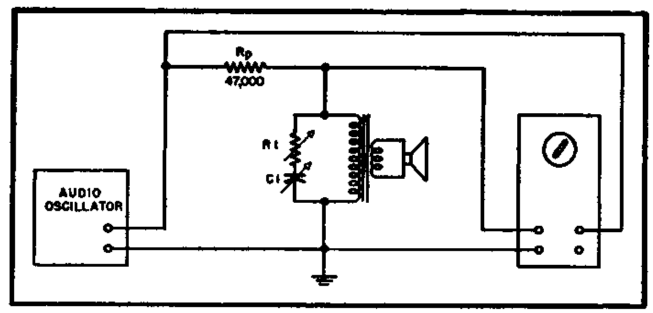

Fig. 3. Circuit used to determine optimum values of R_1 and C_1 of Fig. 1.

The use of direct-coupled voltage feedback eliminates the need for a large blocking capacitor and ensures proper operation of the feedback circuit at the lowest audio frequencies. The current through the feedback resistor R_{f1} must come through the output transformer, which is a disadvantage. Resistors R_{f1} and R_{f2} should preferably be wire wound.

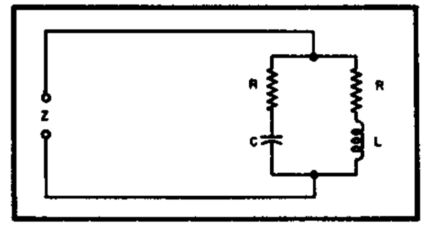

Fig. 2. Series R-C circuit and series R-L circuit in parallel.

Positive current feedback is obtained from a portion of the cathode bias resistor R_{k2} of the 6V6, as shown in *Fig.* 1. R_{k2} is a wire wound potentiometer, and serves as a control to adjust the output impedance. This potentiometer may be replaced by a 200-ohm resistor and the current feedback taken across the full 200 ohms.

The positive feedback control is adjusted in the following manner so that the tube presents zero impedance to its plate load. With the loudspeaker connected, an a.c. voltmeter is connected across the primary of the output transformer with moderate signal applied to the amplifier. A resistor of 5000 to 10,000 ohms (not critical as to value) is then shunted across the transformer primary. The feedback control is adjusted to a point where there is no change in output voltage as this resistor is connected or removed. With full 200 ohms in the feedback control, the output voltage actually increases when the resistor is connected across the transformer, indicating a negative-impedance source. It was found by measurement that the source impedance in the author's amplifier remains zero from 20 to 20,000 cps. It was not checked above this frequency. The drop off below 20 cps is due to the coupling capacitor C_2.

It should be noted that a blocking capacitor cannot be used in series with the negative feedback resistance R_{f1}. If one were used, the negative feedback would become ineffective at some low frequency, yet the positive feedback would still be effective, and low-frequency oscillation or motorboating is likely to occur.

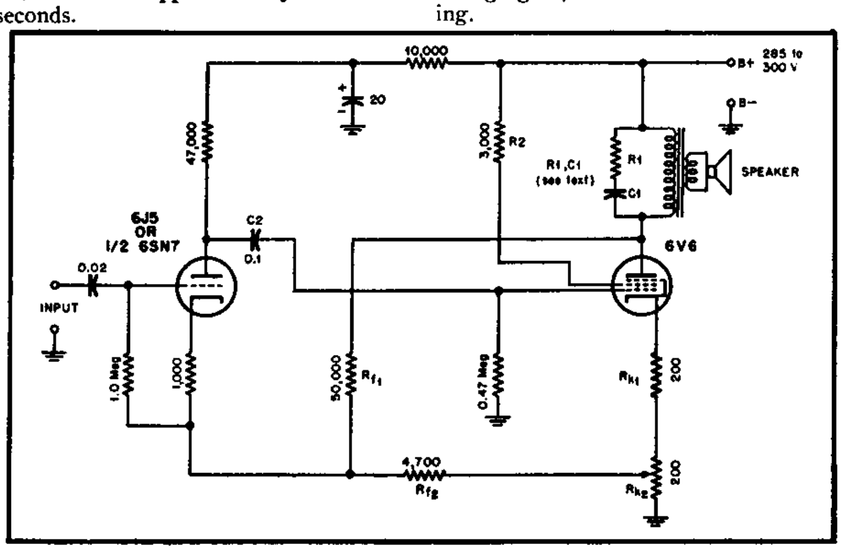

Fig. 1. Schematic of simple two-stage amplifier employing negative voltage feedback an positive current feedback to obtain an output impedance approximating zero.