A sample analysis of a well-known circuit—the Williamson amplifier

By GEORGE FLETCHER COOPER

IN THE two previous articles of this series the general principles of design of amplifiers with negative feedback were discussed. The procedure, essentially, is to design the amplifier, test it, not in the solid but on paper, and then modify the design if necessary to obtain the final circuit.

In this article we shall consider a concrete design, and I shall try to emulate the Butcher, who—

. . . wrote with a pen in each hand
And explained all the while in a popular style
Which the Beaver could well understand.
"The method employed I would gladly explain,
While I have it so clear in my head,
If I had but the time and you had but the brain—
But much yet remains to be said."[1]

The Butcher took three as the subject to reason about but I am going to use instead a high-quality amplifier which has received much attention in Europe and which is, I think, fairly well known in the United States. Before going any further I must state that as far as I know this is a jolly good amplifier and any criticism which may appear is only a reflection of the fact that one designer's meat is another designer's *Poissón*.

The circuit is shown in Fig. 1. The output tubes, type KT66, are closely equivalent to the 6L6, although being British they are rather more powerful or are less conservatively rated. If we neglect the feedback for the moment we can consider this circuit as our preliminary design and we can calculate how much feedback is permissible if the amplifier is to remain stable. The original designer has given us all the stage gains, except for the last stage. Here the total load, plate to plate, is 10,000 ohms, so that the peak voltage across the transformer primary must be 173 volts for 15 watts output ($E^2/R = 15$, $R = 10{,}000$, so that E peak $= E\sqrt{2} = \sqrt{30{,}000} = 173$ volts).

Low-frequency response

There are three primary and two secondary factors governing the low-frequency response. The three primary factors are the two resistance-capacitance interstage couplings and the output transformer. At low frequencies the circuit is completely symmetrical, which makes things rather easier. At high frequencies this is not true, because the stray capacitance at the plate of V2 is in parallel with R7 and the impedance of V2, which is high due to the feedback in the cathode resistor R5.

The stray capacitance at the cathode of V2, a different capacitance, is in parallel with the impedance of V2 acting as a cathode follower. This difference could be quite important if the output stage was operating in class B. It is mentioned here merely as an indication of the special difficulties which the high-frequency response presents when compared with the low-frequency response.

Assuming complete symmetry, the primary factors in the low-frequency response are:

1. C3R8 = C4R9 = 0.05 μf × 0.47 meg = 1/43;
2. C6R10 = C7R11 = 0.25 μf × 0.15 meg = 1/27;
3. L/R in the output transformer circuit. Here R is the resistance produced by the load in parallel with the tube impedance. The load, at the primary side, is 10,000 ohms: reference to data sheets shows that the KT66 has an impedance of 1,250 ohms when connected as a triode. The 6L6 is rather higher, 1,700 ohms, but it has a lower transconductance, so that the main effect of replacing the KT66 with the 6L6 is to reduce the gain without feed-

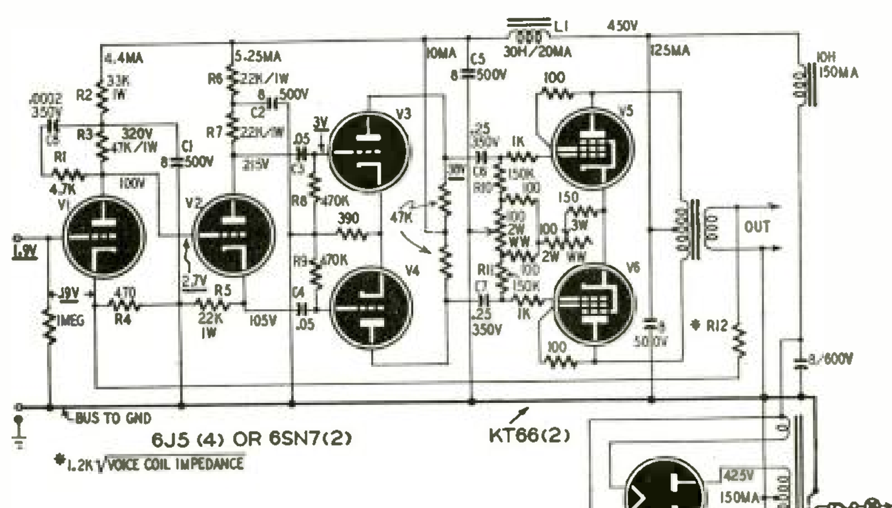
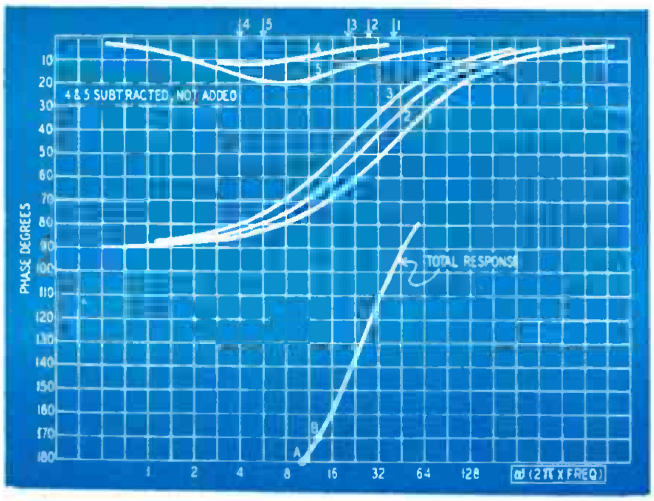


Fig. 1—The complete circuit of the Williamson amplifier. The underlined voltages are the peak signal voltages required for full output of 15 watts.

[1] The Hunting of the Snark, Fit the Fifth, Lewis Carroll



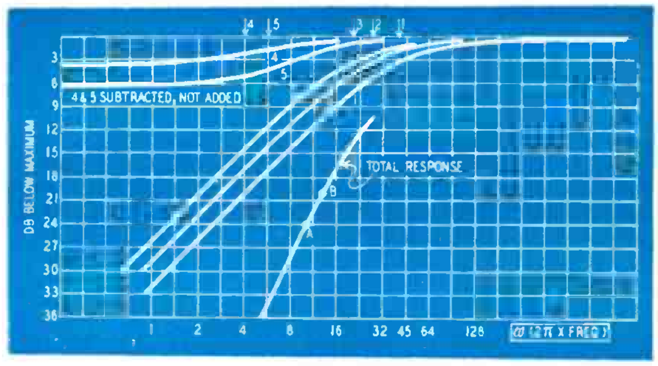


Fig. 2, left—Calculated response curves at the low frequencies of the Williamson amplifier circuit of Fig. 1. Fig. 3, above—The calculated phase characteristic of the amplifier at low frequencies. These curves were plotted by methods described by the author in the second article.

back, and leave the stability about the same as a lower feedback factor.

Two KT66 tubes in series give 2,500 ohms, and this in parallel with 10,000 ohms gives R a value of 2,000 ohms ($1/10{,}000 + 1/2{,}500 = 1/2{,}000$). Using 6L6's, we should have $R = 2{,}500$ ohms.

We assume that $L = 100$ henrics, which gives $L/R = 1/20$. The choice of 100 h may be because this is the largest inductance obtainable with a reasonable size of transformer, or because we want to keep a very good low-frequency characteristic. In this particular amplifier it was chosen because the designer is doing without an air gap and must allow for the increasing permeability at high flux densities.

The L/R factor must, therefore, be free to increase without equalling either of the R-C factors.

The secondary factors are:

4. $C1R2 = 8\ \mu f \times 33{,}000 = 1/4$; $R2/R3 = 33/47 = 0.7$;
5. $C2R6 = 8\ \mu f \times 22{,}000 = 1/5.7$; $R6/R7 = 22/22 = 1.0$.

These secondary factors cause the response to rise at low frequencies, and thus provide a small amount of phase correction. In the critical region this amounts to 30°, and is, in fact, the feature which keeps the amplifier stable.

The response curves

The individual responses are drawn in Figs. 2 and 3, and the total responses are plotted for the critical region. These responses were plotted by the method described in the previous article, and even drawing them rather carefully to please the editor took only about ten minutes. If the figures are examined, we see that we have a 180° phase shift at $\omega = 10.5$, at which point (A on both curves) the amplitude response has dropped by 24 decibels.

If we wish to have 20 decibels of feedback, we must also consider the phase at the point B, $\omega = 13$. This is 170°. Remembering the definition of margins, we see that the phase margin is 10°, and the amplitude margin is 4 decibels (24–20). The reader will see that these margins are rather narrow.

Two other factors must be taken into account in deciding whether they are safe margins. The first, which may not be very large, is the inner feedback loop produced by the choke L1. At the critical frequencies in the region of $\omega = 10$ (about 2 cycles), C5 is a very high impedance, so that V1 and V2 have a common load in L1. This produces a small amount of negative feedback, which I do not propose to calculate.

The second factor is the increase in inductance produced by any signal in the output transformer. The maximum permeability of the core may be five times the initial permeability, and this will shift curve 3 to the left. The reader can confirm, if he wishes, that this does improve the margins. He can also confirm that improved margins can also be obtained by moving curve 1 to the right, by reducing C3 and C4. In general, stability can always be increased by moving the extreme curve away from the others.

One more factor should be noted. At 10 cycles the response without feedback is only 3 db down. This means that we still have 17 db of feedback at 10 cycles, so that the full distortion-reducing effect of the feedback is in force.

High-frequency response

The calculation of the high-frequency response is never very easy because of the lack of essential data. We shall ignore in the first calculation the circuit C8-R1 connected to the plate of V1. The response is then settled by the shunt capacitances of each stage and by the output transformer. Unfortunately the capacitances depend on the way in which the components are arranged, while the transformer's response may be complicated by resonances between the capacitance of one section and the leakage inductance of another.

Let us plunge in boldly, however, and assume a stage capacitance of 20 $\mu\mu$f. We also have the original designer's figure of 30 mh as the maxi-

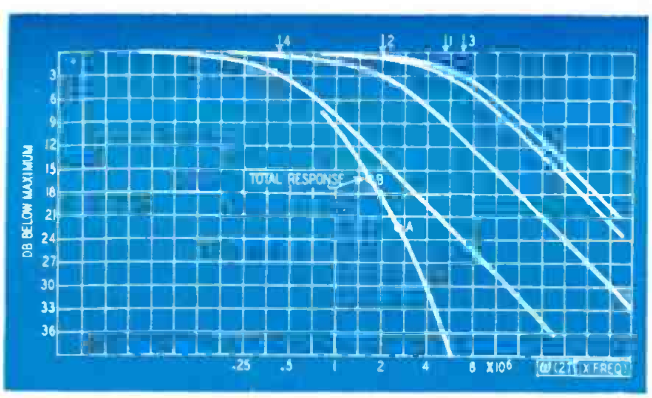
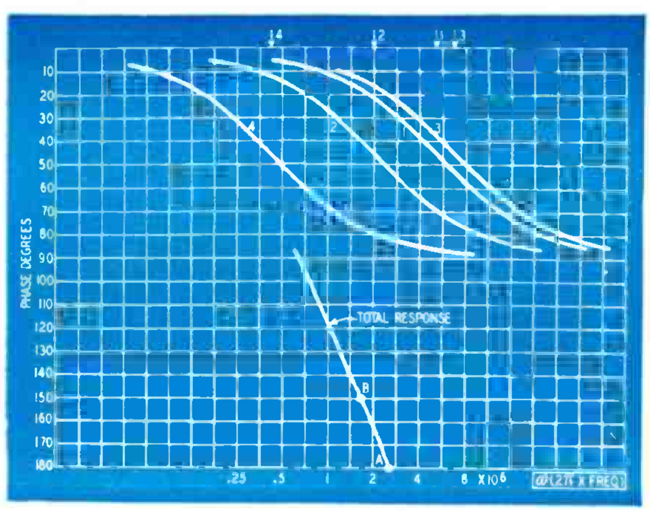
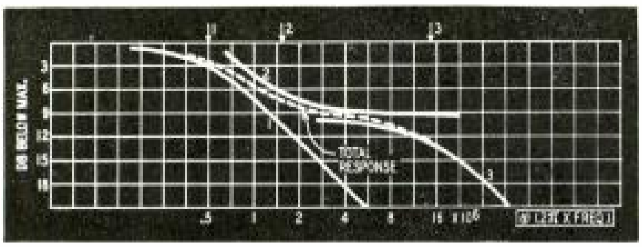


Fig. 4, above—The calculated high-frequency response of the amplifier. At point B the amplitude margin is 6 db. Fig. 5, right—The high-frequency phase characteristic. Maximum feedback for a phase margin of 30° is 16.5 db.





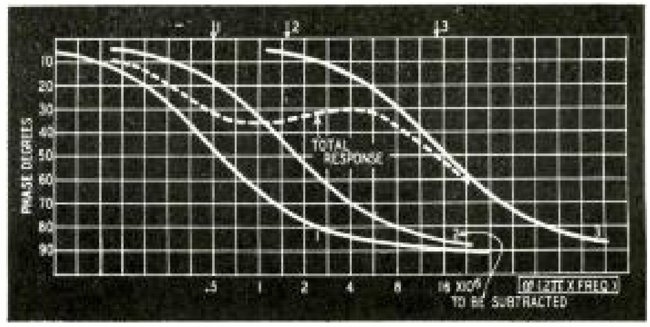


Fig. 6—The high-frequency response of the amplifier including the C8-R1 network. This circuit is added to the amplifier to increase stability at the high frequencies.
Fig. 7—High-frequency phase characteristic with C8-R1.

mum leakage inductance, measured at the primary side of the output transformer. The factors controlling the high-frequency response are then, if $C = 20$ μμf is the plate-ground capacitance and R_{v_n} is the impedance of tube V_n, and allowance is made for local cathode circuit feedback:

1. $C \times R_{v_1} = 20 \times 10^{-12} \times 10{,}000 = 1/5 \times 10^{-6}$;
2. $C \times R7 = 20 \times 10^{-12} \times 22{,}000 = \frac{1}{2} \times 10^{-6}$;
3. $C \times R_{v_3} = 20 \times 10^{-12} \times 7{,}500 = 1/67 \times 10^{-6}$;
4. $L_k/R = 30 \times 10^{-3}/12{,}500 = 1/0.4 \times 10^{-6}$.

These factors give the curves which are shown in Figs. 4 and 5. These were drawn in just the same way as before, using the simple templates, and only the important part of the total response characteristic has been drawn. The phase shift reaches 180° at $\omega = 2.6 \times 10^6$ (f = 300 kc). At this point the amplitude characteristic has fallen by 22.5 db, indicated by the point A in Fig. 4. If we take 150° as the safe limit, we have B, and a maximum feedback is of 16.5 db. The amplitude margin is then 6 db, and the phase margin 30°.

Increasing stability

One way of increasing the stability is to increase the leakage inductance; another is to reduce the stray capacitances, especially that of the first stage. The reader will do well to recalculate these curves for, say, 50-mh leakage inductance and 15-μμf capacitance. In the original version of this amplifier it is clear that the margins were rather small for the use of production transformers, for the circuit C8-R1 has been added. Let us see what this does.

The capacitance C8 is 200 μμf. At a frequency $\omega = 1/C8 \times R_{v_1}$, the response of the first stage will start to drop, and it will run down to meet a curve defined by C8 and R1. At still higher frequencies the response will drop owing to the 20-μμf plate capacitance in parallel with R_{v_1} and R1. Instead of the curves 1 in Figs. 4 and 5 we will have the curves shown in Figs. 6 and 7. We need the characteristic factors:

$1/\omega_1 = C8 \times R_{v_1} = 200 \times 10^{-12} \times 10{,}000 = 1/0.5 \times 10^{-6}$;

$1/\omega_2 = C8$ (R1 and R_{v_1} in parallel); $= 200 \times 10^{-12} \times 3{,}000 = 1/1.5 \times 10^{-6}$;

$1/\omega_3 = C$ (R1 and R_{v_1} in parallel); $= 20 \times 10^{-12} \times 3{,}000 = 1/15 \times 10^{-6}$.

We could now redraw Figs. 4 and 5, but this would take up too much space for this article, and it is sufficient if we simply compare the curves 1 of Figs. 4 and 5 with the total response curves of Figs. 6 and 7. At $\omega = 2 \times 10^6$, for example, we had a contribution of about 1 db and 20° from the simple circuit, and the addition of C8-R1 has increased the attenuation to 7.5 db and the phase to 32°.

This means that the phase is now just over 180° at this point, and the attenuation is about 26 db. The amplitude margin of 6 db will then allow us to use 20 db of feedback. At $\omega = 1.4 \times 10^6$, the C8-R1 circuit gives us 6 db and 35° instead of 0.5 db and 15°, so that the total response at this point will have a phase shift of 160° and will be 19 db down.

By examining a few more points we can determine the phase margin exactly, but it is a little under 20°. These margins are rather tight; but, as we are making no allowance for the output transformer capacitance and as any assumed capacitance can be in error by ±25% or more, we must not be too critical. In a later article we shall see how to deal with high-frequency instability.

At this point let us look back. We have taken as a design basis the circuit shown in Fig. 1 and have made certain assumptions which have enabled us to draw the amplitude and phase characteristics. These, in turn, showed us that we could apply 20 db of feedback without low-frequency instability, but that we require the stabilizing circuit C8-R1 if the amplifier is not to be unstable at high frequencies. We can also see that, without feedback the response being only 3 db down at 10 cycles, we get the full feedback over this range for the reduction of distortion and intermodulation.

The feedback circuit

One more thing remains to be determined. In the actual design process we must calculate the value of R12 which will give 20-db feedback. Usually, of course, we must just calculate the gain without feedback, but it is assumed that the reader knows how to do this. The designer tells us, or your own calculations will tell you, that the input voltage between grid and cathode for 15 watts output must be 0.19.

We shall ignore the local feedback produced by R4 and assume that with R12 connected we want the gain to drop 20 decibels, making the new input for 15 watts output 1.9. Then we have 1.9 volts from grid to ground, 1.71 volts from cathode to ground, and the necessary 0.19 volt from grid to cathode.

Let us assume that the transformer is designed for a 3.6-ohm secondary load. The 15 watts output then corresponds to $\sqrt{3.6 \times 15}$ volts across the load, or 7.4 volts. My calculations give R12 = 1,570 ohms to produce this required 1.71 volts at the cathode, while the original designer gives 2,200 ohms.

The reason for this discrepancy is the difference in what is meant by 20-db feedback when the main feedback loop also involves a local feedback of 6 db. Two different answers are obtained depending on whether the feedback is removed by disconnecting R12 or by short-circuiting R4 to alternating current with a very large electrolytic capacitor.

In commercial design one more factor needs to be considered. Is the amplifier open-circuit stable? Often we need to have an amplifier switched on, but idle, and, if it operates from a common supply system with other amplifiers, it cannot be allowed to be unstable even when not in use. To test this we must redraw the characteristics for the amplifier with no load on the output transformer. The general question of load impedance will be discussed in a later article.

These calculated response curves are, of course, not the same as the actual measured response curves of the amplifier. We cannot, without a great deal of cumbersome mathematics, account for such things as tolerances of the components, stray wiring capacitance, and a number of other factors. However, most of these items are rather small, and they also tend to average each other out. That is, the tolerances may be either plus or minus.

What we do get from these curves is a very substantial idea of how the amplifier will behave once it is constructed. We immediately see any important flaws in the basic design so that the necessary corrections can be made at no cost of time or parts.

The next article will describe a loudspeaker amplifier designed by the writer. Unlike Mr. Williamson's amplifier, the design is based on a minimum size of transformer, and a comparison of the two designs will show the reader how flexible the design method is in some ways, and how inflexible are some of the restrictions.