Hi-Gain Phase-Splitter

Original Title:
Push-Pull Phase-Splitter: New High-Gain Circuit

From a Wireless World article Aug 1947, by E. Jeffery A.M.I.E.E

An Open-Loop Signal Gain Of 1,200 Times And A Perfectly Symmetrical Push-Pull Output From Just Two Valves!

THE advantages of RC-coupled input and phase-splitter circuits for push-pull amplifiers are now well recognized, a wide variety of circuits for this purpose having already been fully evolved during the 'past decade' [= 1930's to 1940's!]. If these circuits are investigated it is found that they possess varying degrees of merit in producing symmetrical output signal voltages, but share the common disadvantage of inherently low gain. (To clarify see article[1], 'Negative Feedback In Amplifiers', Fig. 3(a) and associated discussion text, as a reference to this particular circuit type.)

The circuit described here, however, will be shown to possess a very high degree of symmetry with a large amplification, using only two valves.

The type of cathode-follower phase-splitter shown in Fig. 1. is perhaps the most generally used type [or should be], because of its simplicity and the high degree of symmetrical balance obtainable between the outputs Eo and Eo2, being dependent only on the accuracy of the values of RL and Rc. [Also it is self-biasing with a capacitor input, which means that the precise DC anode voltage of the preceding stage is not critical, so that the input stage is free to be set up for the best DC working conditions.] Of course it inherits the disadvantage of a cathode-follower in that its voltage gain is slightly less than unity (refer to article[2], 'Input Impedance Of Cathode-Follower Phase-Splitter' for a discussion).

If we now consider this arrangement to be preceded by a pentode stage, then the overall gain is virtually that of the pentode alone, typically about >=100 times [EF36-37-86 type].

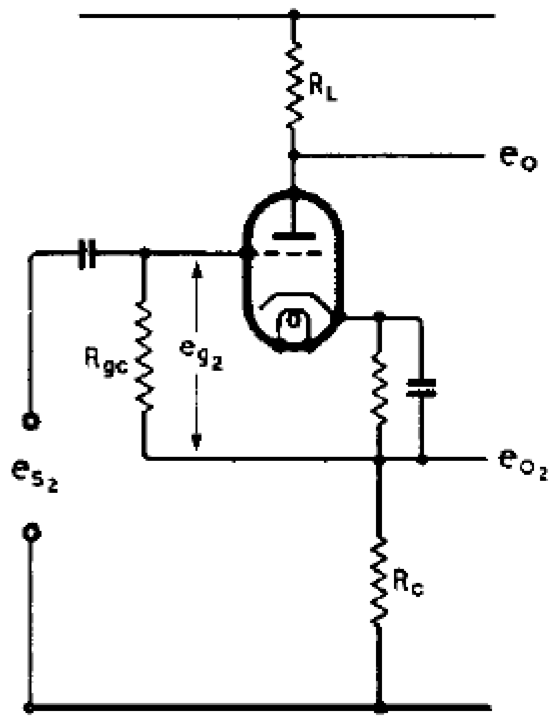

Fig. 1. Cathode-follower phase-splitter

It is well-known that the input impedance of Fig. 1. can be extremely high – depending to some degree on the specific valve type employed – and typically may be considered to be 10 x Rgc. If Rgc is made 250kΩ, then the input impedance can be assumed to be 2.5MΩ. (**Note**, however, that this is true *only* if the cathode bias resistor is shunted with a capacitor as shown, so that the bottom end of Rgc is directly AC coupled to the cathode.) This does not appreciably shunt the AC anode load impedance of the pentode, hence the gain of the pentode is determined almost entirely by the value of its anode resistor alone.

The significant point being, that this large input impedance value for Fig. 1. *matches* the internal anode resistance (Ri) of an EF36-86, which is also 2.5MΩ. If it were possible to contrive that the entire AC anode load impedance as seen by the pentode were this value, then the gain would be huge, solving the low gain problem of this simplest configuration of a cathode-follower phase-splitter sourced from a single input valve. [The simplest configuration with the least number of valves is to be preferred, in order to reduce phase shift problems in a closed-loop amplifier using negative feedback (see article[1].)]

High-Gain Circuit

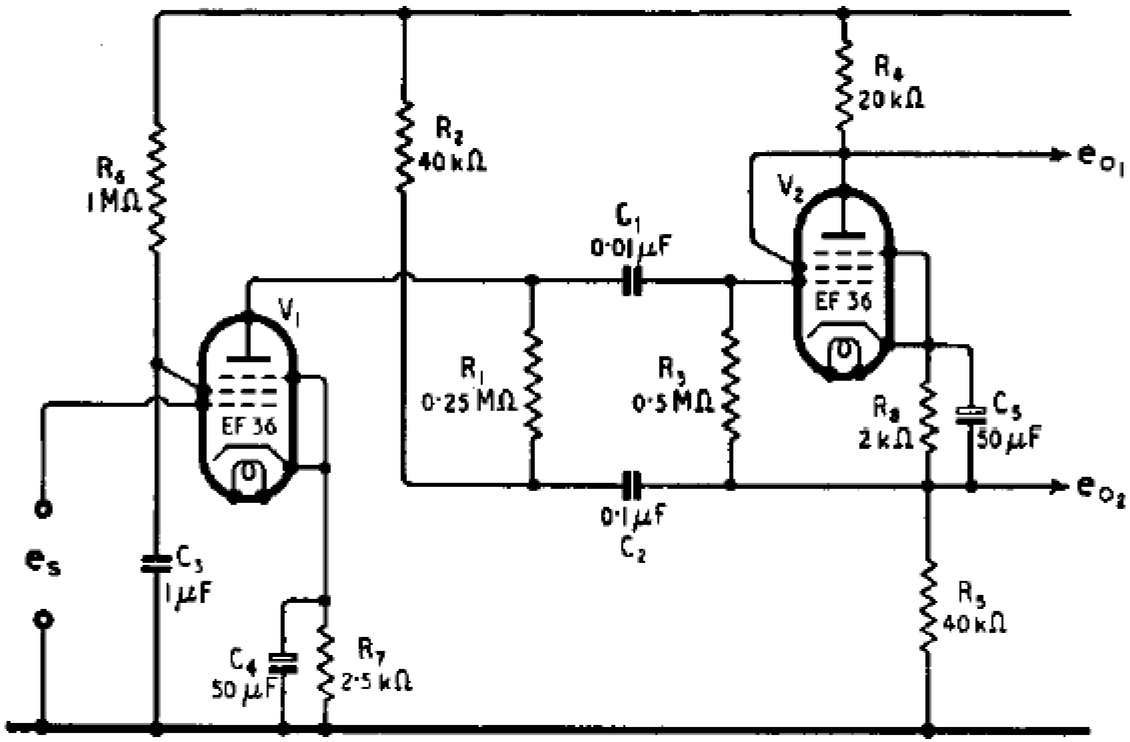

Fig. 2. Practical form of the high-gain phase-splitter and preceding pentode stage.

This can be achieved in practice by applying the input impedance multiplying effect of the cathode-follower to *both* Rgc *and* the preceding pentode's series anode resistor, by the arrangement shown in Fig. 2. above. [This is recognizable as the old 'boot-strapping' technique, once popularly employed in complimentary push-pull transistor amplifiers. (In that case it was used to ensure that the voltage drop across the bias resistor chain, for the complimentary output transistors pair, remained constant and not vary with signal amplitude, which could upset the carefully set quiescent current for the output stage and thereby introduce the dreaded 'zero-crossing' distortion.)]

The output from Eo2 is returned to the positive end of R1, the anode resistor for V1, via C2, such that the AC voltage at this point is the same as that at the anode end of R1 minus the grid to cathode voltage difference of V2. The effective AC impedance of R1 may therefore also be assumed to be 10 x R1, and so V1 anode sees a total effective AC impedance of R1 and R3 in parallel, as usual, but x 10.

To be more accurate we need to find the actual impedance of the second stage using the procedure described in article[2], but where R1 is in parallel with Rgc (R3). Mr. Jeffery calculated R1 and R3 in parallel as 168k, in actual fact it works out more accurately to be 166.67k (see the parallel resistors calculator at bottom of article[3], 'Calculating Gain Of Valves'). Where V2 is an EF36-86 wired as a triode, μ = 28 and Ri = 10kΩ, and using the input impedance calculator of article[2], R_{in} (the input impedance of the cathode-follower phase-splitter stage V2) works out to be 2.03MΩ. Mr. Jeffery calculated it to be 2.05MΩ, which it would be if R1 // R3 = 168k. He was also quite convinced that the EF36 amplification factor (μ) ought to be as much as 4,500.

The amplification of V1 would then given by:

$$\mu \frac{2.03\text{M}\Omega}{2.5\text{M} + 250\text{k}} = 3{,}321.82$$

if μ = 4,500, which is unlikely in practice. Even so Mr. Jeffery must have been mistaken in getting a result of 2,030 from a R_{in} of 2.05MΩ, I couldn't duplicate this even while following the equation step-by-step and double-checking. I always got a value of 3,354.55 instead. (Of course he didn't have the advantage of modern calculators and computers! Instead he had to laboriously work it out 'by hand' on paper.) In any case, in reality, a μ of about 1800 would be typical for the average EF86 or EF36, whereas a 'good' EF37 might run to 2200.

You can use the valve gain calculator at the end of article[3]. For μ = 1,800 and R_{in} (RL) = 2.03M, R1 (Ra) = 250k, the gain is 1,328.73 times. We must then multiply this by 0.9, the 'gain' of the V2 stage, which gives about 1,200, and in fact a table near the end of Mr. Jeffery's article gives 1,210 of actual measurable gain for a supply of 400V. So there!

The AC signal level balance error between the two phase-splitter outputs Eo1 and Eo2 is quoted as being within 1.2%, and that's using the grotty old 10% resistors they used to have in those days. Using modern close-tolerance types, the unbalance can be regarded as completely negligible. It must be noted, however, that R2 & R5 are each <u>exactly twice the value</u> of R4, this is because R5 is shunted by R2 for AC, in other words Eo2 has an AC load of 20 kilohms, exactly the same as R4 for Eo1. Further additional loading by the grid bias resistors of the output stage should also be equal. Using currently available standard values, R2 & R5 can be 47k, making R4 23.5k, which is most easily achieved with 2 x 47k in parallel.

Design Considerations

The circuit was specifically developed to drive valves of the PX25 class [= KT66 as triode, see below], and the values were therefore chosen to give a large peak output rather than maximum gain. If it is required to drive valves requiring a smaller grid swing no doubt much higher values of gain can be achieved. [Gain may be increased even further by raising the value of R3.] It should be pointed out to those evolving their own designs that the dynamic load on V1 is very much greater than the DC load, while the AC load on V2 is *less* than the DC load.

For smaller output voltages, say up to 25V r.m.s. at each output point Eo1 and Eo2, the component values are not critical, while other types of [output] valve have been substituted with only minor changes circuit changes, e.g. bias and screen resistors.

It is of interest to note that the EF36 strapped as a triode for V2 gave better linearity than any of the triode types investigated. There is a (small) chance that the large heater-cathode voltage difference might be a problem, if so, the heater current for V2 should be derived from a separate, isolated, 6.3Vac source, whose centre-tap is connected to a potential divider between HT and 0V, such that the voltage at that point is within ±100V of that on the cathode. The centre-tap point must be decoupled to 0V with an electrolytic to ensure hum rejection.

In order to avoid loss of gain, it is essential to bypass the cathode-bias resistors of each stage. [Here 47μF will subsitute the 50μF values shown in Fig. 2.]

Note that the value of C2 must not be too large, else LF instability may result. I tried something like this before a little while ago, in that case the capacitor was 1μF, and what I got was a 1 Hz oscillator! In fact the low-ish value shown for C2 will roll-off the gain at the low frequency end of the audio band, bearing in mind that it is AC loaded by the value of R2, and therefore ensure LF stability.

Applications Of The Circuit

Mr. Jeffery writes: "The circuit has been used over a number of years in a wide variety of amplifiers and has proved to be remarkably stable and free from undesirable traits." The preferred output stage should use KT66 valves strapped as triodes, this method of connection giving a performance equivalent to that of the PX25 type, with the added advantages of shorter grid swing, indirectly-heated cathodes and Octal base. Output power should be about 15W r.m.s. using output transformer VT1041, see the amplifier write-up about 'Trident-66D' and associated circuit PDF for a working example.

Negative feedback is applied to V1 by inserting a 10 – 100Ω resistor between the bottom ends of C4 & R7 and the 0V rail, then connecting the feedback resistor to the same (upper) junction. The signal voltage gain between input and the transformer speaker output secondary is then the ratio of these two resistors, minus about 10 – 12mV, which is the input sensitivity of the whole amplifier for maximum output without feedback. The feedback resistor should not need a HF bypass capacitor to be added.

The negative feedback reduces the distortion to 0.5% at maximum output, while the output impedance, for the purpose of loudspeaker electrical damping (see article[1]), becomes $120/n^2$, where 'n' is the turns ratio of the output transformer primary. For a VT1041 the turns ratio is about 28:1, making 'n' 28, and hence giving an output resistance of about 0.15Ω.

Where more output power is required, the KT66's may be operated as tetrodes or 'ultra-linear', or KT88's may be used. However Mr. Jeffery prefers paralleling two pairs of push-pull KT66's wired as triodes to obtain a larger output and retain quality.

"No originality is claimed for this circuit, as a search of the literature has shown that it appeared in the USA some years ago [1930's]; however, so far as the author [Jeffery] is aware no analysis of the circuit has previously been published."