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Notes on a Work in Progress

The PEARL SC 280.1—January, '92.

THE FOLLOWING is a brief update on the SC280 power amp project. I have given the amplifier a complete re-work since the last '280 Audio Note was written and have a much better collection of circuits to assemble into a finished whole. I am going to write a full update to the original 'Note, but for the time being I'll give a brief overview of the new design.

I decided that I was trying to make the original UL cascode front-end circuit do too many things. I wound up paying a price by so doing in that I was forced to capacitively couple to the output stage and suffered from somewhat low gain as well.

The new front-end stages are much better and are capacitively coupled into low output-impedance driver-stages that include controls to move the operating point of the output-stage from lean Class B to full Class A and allow the standing current to be easily balanced. I have also re-designed and improved the current source that is a vital part of the diff-amp configuration.

The output of an ideal differential amplifier is a pair of perfectly phase and amplitude matched signals that are, respectively, exactly in phase and out of phase with only the difference between the signals deliberately impressed upon the two input grids (gates or bases).

Despite its name, a differential amplifier does not inherently provide anything even approaching such ideal performance and the situation is aggravated when the amplifier is driven in single-ended fashion—with one of the input grids tied to signal ground. Rather than go through a lengthy explanation I have simply included rebuilt editions of some of papers on diff-amp operation. Careful study of these will give you the information needed to understand the approach taken here.

In the text that follows I am assuming that you have read and absorbed the articles included herewith.

Significant even-order distortion generation is usually seen in diff-amps executed with a simple resistor as the common-cathode element. This happens because the degeneration signal created across this resistor by current flow in the "driven" tube winds up being anti-phase with—rather than in-phase with—the signal flow created through the cathode drive to the "undriven" tube. By this mechanism even-order products are created that manifest in the output of the amplifying stage. The usual approach to this problem is the replacement of the resistor with a degenerative current-control device that acts as a virtual high-value resistor.

An active current-source will usually improve the balance between the two output voltages but not reduce distortion in the outputs very much.

The function of such a current-source is:

- to provide the required cathode drive to the side of the diff-amp that is not driven when the amp is run from a single-ended source
- to provide any correction signal required to drive the common cathode point in such a way as to ensure that the outputs of the amplifier remain balanced in the presence of any unbalance in a push-pull drive signal from a balanced source such as a line amplifier.

When a diff-amp is driven from a single-ended or unbalanced source the undriven side has its grid tied to signal ground, so its cathode must be driven to provide the required change in grid-to-cathode potential required for signal amplification. Without going into yet more theory, suffice it to say that the even-order generation mechanism creates products that can be reduced in magnitude by the application of feedback that improves the common-mode performance of the stage.

The hybrid current source included in this design has been designed to have such an effect and its basic operation is explained quite well by J.C.S. Richards in "*An Improved Type of Differential Amplfier*".

The basic idea is to create a circuit that will provide a drive signal to the common-cathode connection of the diff-amp that is of such an amplitude and phase as to reduce the creation of even-order products and/or reduce the output of the stage when its two input grids are tied together and driven with respect to ground by the same signal voltage. This is known as driving the amplifier with a common-mode signal and the resultant output—from diff-amps with less than perfect common-mode rejection—causes the two outputs to move in-phase with each other, either towards or away from signal ground. The zero-signal position of any reproduced waveform is altered by this common-mode component with undesirable effects on all the of the subtle yet musically vital information buried at the zero-crossing point.

A typical active current-source exhibits a very high output-impedance that is a natural outcome of both its purpose and design. When driving a low impedance point such as the common cathode connection of a diff-amp, each cathode's capacitance to the grounded filament presents an undesirable load to the high impedance point. Since a filament supply with low capacitive coupling to ground is a very difficult thing to manage, I opted to design a current-source with a low output impedance. Since a truly high-output-impedance current source must basically consist of a high gain amplifier, all that was required was to place an active load "on top of" the current source and take the output from the low-impedance output-point of the active load.

The Driver Stage

Each side of the output stage is driven by a pair of conventional White cathode followers. Each follower is directly coupled to half of the tubes in one side of the output stage and both followers are capacitively coupled to one side of the front end.

There are effectively three bias adjustments; one allows the operating point of both sides of the amplifier to be changed simultaneously while the other two facilitate an easy re-balance of the standing current in each of the push-pull groups of the output stage. For the White followers, I use tubes from the 5687 family which includes the 7044 and 7119 (not 7199). These are high-transconductance devices that show extremely good linearity. For those octal fans out there, the 6SN7/5692 has almost the linearity at about one third the g_m. There are several other low μ tubes that will also do the job; for a couple of very high peak-current types look up 6BL7 & 6BX7.

For this scheme to work well the use of output tube-pairs or -triplets that are fairly well matched in terms of their standing current vs. grid bias is necessary.

While this may seem a bit odd, the fact is that any differential or push-pull amplifier should be built with devices that are as closely matched for AC characteristics as possible. Otherwise the feedback loop is doing more work than necessary to compensate for unequal side-to-side gains. Given that the AC characteristics should be matched, it's no big deal to match for the DC characteristic (standing current vs. grid bias) at the same time. A little effort in the tube matching department is well worth it.

The Output Stage

This is the stage in the amp that seen the fewest changes. All that has been done is to eliminate the individual bias arrangement previously connected to each tube. Along with this went the bias-point adjustment adjustment with both of these being fitted into the driver stage as previously mentioned. The reason for going to a directly-coupled drive to the output-stage grids is that I recently discovered that all beam tetrodes seen to date—KT77, KT88 (all versions- MO Valve, Richardson(Cetron) & Chinese), KT90A, 6550, 6CA7, 6L6, 807, 8417, 5881—begin to exhibit substantial positive grid-current (electrons flowing out of the grid) at grid-voltages as far negative as –10 volts. At zero grid-volts, grid-currents of several mA. are commonly seen. These currents wreak havoc with the bias point stability of conventional high-impedance capacitively-coupled circuits where large value resistors—100KΩ or larger—are used to tie individual output-tube grids to ground.

A driver stage used to run these tubes in any form of tetrode operation—UL, unity coupled, straight-ahead fixed-screen tetrode, etc.—must be able to sink significant current flows. A side benefit accrues from the satisfaction of this requirement by the consequent ability of the driver stage to sink grid-current flow from the output tube grids when they are deliberately driven positive, as when handling high-power peaks. This gives the amplifier a very authoritative sound because clipping is handled graciously and occurs at power levels substantially in excess of those available when the output stage is only capacitively coupled.

Speaking of output power, the choice of output tubes is obviously pivotal. I have recently finished an in-depth survey and evaluation of all the presently available devices. Of the lot, only two are worth serious consideration; the Tesla EL34 and the EI/VTL KT90A. The Tesla part is made in Czechoslovakia by Tesla while the KT90A is made for VTL in Yugoslavia by Electronska Industrija.

Not to dwell upon the recent brouhaha in *Stereophile,* the KT90A is alive and very well. While there were initial problems, these have been solved and David Manley is owed a heart-felt "Thank you!" from all of us for having the…uh…spheroids to risk the enormous sum required to start production of a new output tube. My small contribution was to re-design the plate structure of the tube to linearize the design and sort out the less-than-ideal response of the early devices. I have produced and included a full spec sheet that includes curve tracer photos of the plate curves in both triode and tetrode configuration.

The Tesla EL34 is simply an excellent device and easily the best of the Eastern European EL34 production. Unlike the rest of the Continental parts, it is rated for operation at 800 plate volts and 400 screen volts. This is the original Mullard spec and one that all EL34s should meet but don't; the Chinese tubes will go up in a flash of blue light if you run them at 800 plate volts. I've recommended the Tesla to a number of manufacturers and to a man they have all made the switch with the likes of Marshall (US and UK) and VTL enjoying good success

Matched KT90As cost $68.00US /pr. and matched EL34CBs are $28.00US/pr; shipping is extra. PEARL power-tube coolers are $11.00US ea.

POWER SUPPLIES

I have included basic schematics for the power supply and regulator circuitry. Although these give the essentials there is still a bit of fine tuning to be done.

The soft-start for the B+ may look a bit primitive but the action of this circuit is simplicity itself. An inrush-current–limiting thermistor, an LED or incandescent pilot light and center-off switch that makes momentarily in one direction from center and continuously from the other are all the parts needed. Substantial currents must be handled by this circuit during normal operation so nothing less than a stout parallel-wired DPDT switch will do the job properly.

The way I set this up is to mount the pilot light vertically above the switch (on a vertical control panel) and label the three switch positions as follows:

- Up position – RUN
- Center position – OFF
- Bottom position – START

Operation is easy, switch to START, wait until the pilot light comes to almost full brilliance and switch to RUN. Note that a neon indicator won't turn on gradually like an LED or incandescent one will.

Please don't get carried away and design an automatic soft-start using relays or SCR's to switch the AC.

Unless they are switched at the zero-power point of the AC waveform, SCR's create switching noise and this is more trouble than it's worth to design around.

Relays exhibit a couple of problems caused by oxidized contacts that are likewise hard to deal with.

The most commonly used sort of power supply filtering configuration is a single, large capacitor placed across the full-wave rectified AC. Without going into great depth, it's enough to say that the replenishing currents required to bring the capacitor back up to voltage after a period of discharge can be many times the average current drawn as quasi-DC from the supply. This is problematic because the peak charging currents can run the core of the power transformer in hard saturation.

Saturation of the core starts a self-stoking vicious cycle because iron that is saturated effectively drops out of the magnetic circuit. Only the reactive value of the primary winding as an air-core inductor in parallel with the reflected load from the secondary—effectively a momentary short circuit—remains to hold off the enormous current flow that line voltage will force through the primary winding; thereby driving the core into harder saturation. Once the full-wave output from the bridge has peaked, the current inrush drops to back to the idle value and everything starts to settle down until the bridge's full-wave output comes up once again to meet the decreasing voltage at the capacitor. This sort of activity is hard on relay contacts.

All relay contacts oxidize over their useful life and such oxides are poor conductors. When large currents are pulled thorough these unavoidably non-linear resistors noise must be generated. Being of a high frequency nature it will ride right through the stray capacitive coupling between a conventionally wound power transformer's primary and secondary windings and through the effective shunt capacitance across the diodes in the rectifier bridge. It sees a high impedance to ground at the filter cap because of the cap's significant stray inductive reactance at the frequencies I'm discussing and winds up on the power supply hot *and ground* rails. Note that bypassing the main filter cap with a foil cap only provides a low-Z path through which high-frequency noise-currents can flow, the ground bus is still affected by the HF noise

Relays that are placed in the circuit so that AC energizes the pull-down coil create minute vibrations in their contacts at twice the line frequency. These are caused by the modulation of the magnetic flux caused by the AC used to energize the coil. This chatter is, of course, transferred to the contacts; combined with the effects mentioned above, the sonic outcome is much less than pleasing

I recently had the opportunity to listen to a very "hashy" sounding amplifier that used such a circuit. When the relay was removed from the circuit and the power supply allowed to "thump-start" the sound was much improved!

FEEDBACK OPTIONS

I am working on a number of feedback options, two of which are shown in the schematics provided.

The 6AU6 circuit uses conventional negative, voltage feedback but I am applying it to the screens of the front-end tubes rather than into their cathodes. This Ultra-Linear operation has a number of advantages I'll discuss later but will say now that the output Z of the stage is much reduced and distortion and bandwidth are very good.

The 12AX7 circuit is for those who want to try a circuit using no loop feedback. As with the 6AU6 front end, there is common-mode feedback within the diff-amp and cathode feedback in the output stage.

Another circuit is in the works using a triode output-stage with positive, voltage feedback applied to the cathodes of the output tubes and negative, voltage feedback applied to 6AU6 front end as above. If you stop and think about how triodes and pentodes work, this promises to be quite interesting.

It's a seldom appreciated fact that a triode is unique amoung amplifying devices in that it provides a fair measure of internal feedback. For completeness, I've included a rebuild of "*Inherent Feedback in Triodes*" by H. Stockman, SD, for you to wade through but the concept is simple.

In a nutshell here it is: the output of almost any triode voltage amplifying circuit appears between AC ground and the junction of the plate and the plate load-resistor. Note that the other end of this resistor is tied to *AC ground* by the power supply capacitor, so the output voltage appears across the load resistor and is a function of current flow through that resistor.

A triode's plate current is, all other things remaining constant, a function of plate voltage. When the control grid is made more positive with respect to (w.r.t.) the cathode, more electrons are allowed to flow to the positively charged plate that attracts them. The strength of this attraction is a non-linear function of the plate-to-cathode voltage.

With an increase in plate current comes a greater voltage drop across the plate load resistor and a decreased voltage drop across the tube. The outcome is that the plate becomes less positive w.r.t. the cathode and therefore exerts less attractive force on electrons trying to pass through the mitigating negative-field at the grid. This reduced attraction results in lowered plate current, reduced voltage drop across the plate load resistor and less output voltage from the stage.

This is a negative feedback action that could be called *plate degeneration* and it's responsible for the lowered power-output of triode- vs. tetrode-connected tetrodes. It can be offset and the amplifier turned into an oscillator by the application of a positive—aiding—feedback voltage into cathode.

On its way to becoming an oscillator the circuit passes through a region of operation that goes from triode through pentode. It's interesting to note that variable-tap operation of an Ultra-Linear output stage can result in an initially tetrode-connected tube moving to triode connection through the application of negative feedback to the screen grid.

The so-far-overlooked outcome of all this is that you can create the transfer-function-equivalent of an Ultra-Linearly connected tetrode by applying a small amount of positive feedback to the cathode of the same tube connected as a triode—screen grid tied to the plate. This will not cause the amplifier to oscillate if the amount of positive feedback is kept low; +3 to +5dB. The use of positive feedback has several advantages that can be meaningfully exploited.

Before getting into this it's probably worthwhile to review some of the advantages of UL operation of the output stage.

Briefly, the transfer curve for a power pentode is a curved line and for a triode it's also a curve but with the "bow" going in the opposite direction. Within the middle ground lies an area where an ideal straight line could be created if you did the right thing. A clever fellow, Blumlein, figured that a little negative feedback into the screen grid would achieve the desired result and derived his feedback signal from a taps on the half-primaries of the output transformer. From this work in the 1930s, the mode of output stage operation later dubbed "Ultra-Linear" by Hafler and Keroes was born.

Negative FeedBack normally reduces the effective output impedance of a stage and that is the case here too. Although the effect is dependent upon the Z of the driven load, it doesn't take much NFB to reduce the output Z of a pentode stage by an order of magnitude or more.

UL operation provides enough NFB to achieve very triode-like output Z and there isn't any real advantage to going to triode operation to achieve yet lower Z.

One of the effects of positive feedback is that it creates a situation where the amplifier is much more capable of "swallowing" back EMFs, such as those created by resonating bass speakers. While the amp must be able to "get a good look" at the load—there can't be any significant series DC resistance interposed—the sonic effect can be very gratifying.

I'll work on this later in the year and by the winter should have something definite to say about *ENHANCED–TRIODE-OPERATION*.

If you reverse the cathode coupling connection of the output stage to achieve positive, rather than negative, feedback, you may get too much and have a high-power oscillator on your hands for a short time. I expect that I'll have to place special taps in the secondary to facilitate this mode of operation. My recommendation is that you let me fiddle with this first and if it works as well as I think it will an update will be available.

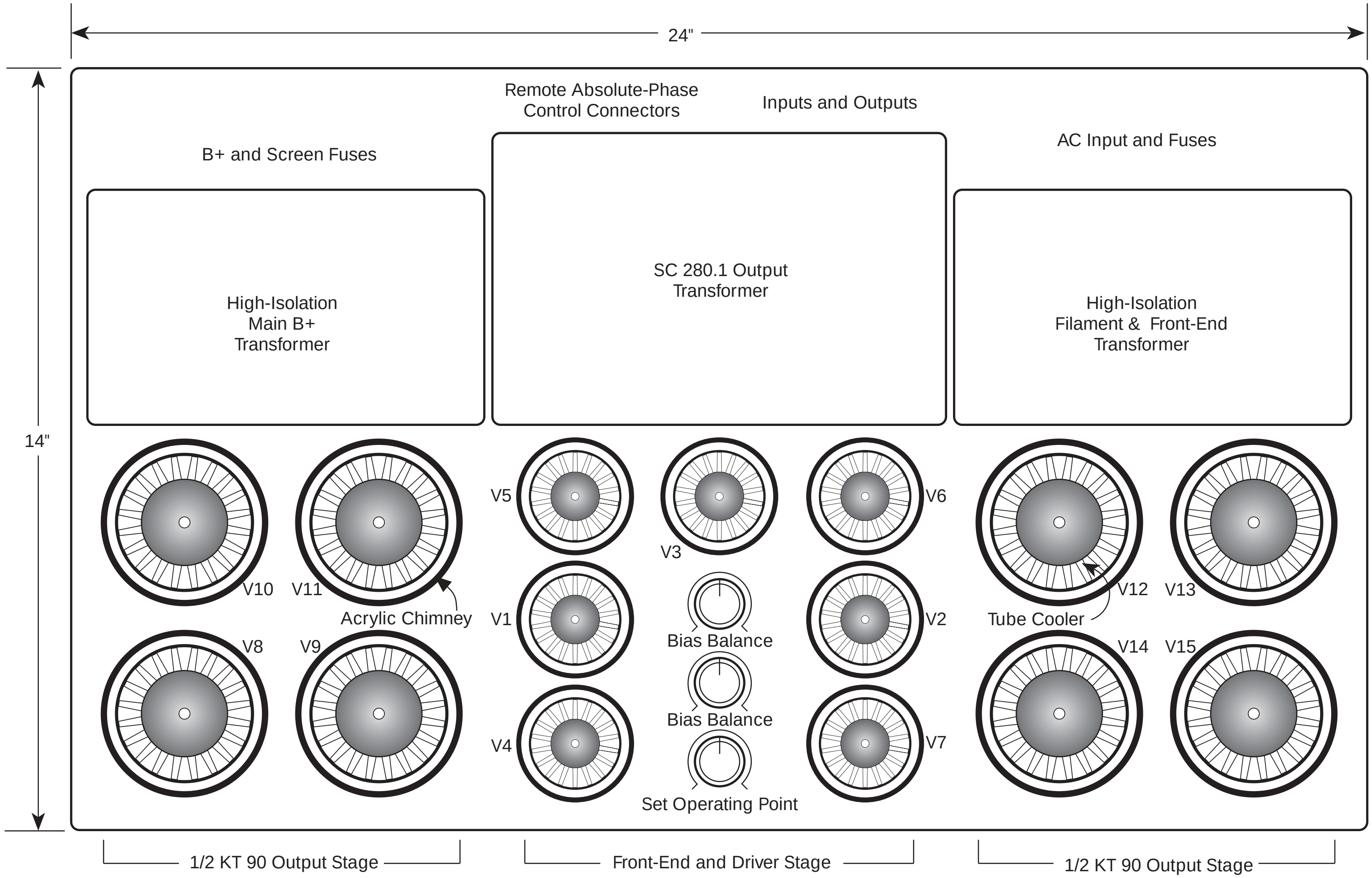


Fig. 1. The PEARL SC 280.1 monoblock is shown in plan view. Note the symmetry of the layout. Fans within the chassis force cooling air up through all of the cooler-chimneys resulting in the sort of tube cooling performance shown in Fig. 1 of *Audio Note* 23.1. All power-utility AC is contained within sealed, bulkheaded compartments directly below the high-isolation power transformers. Only isolated, filtered and initially rectified AC is allowed into the rest of the chassis.

The remotely-controlled absolute-phase-switching circuitry is operated from an umbilical that runs from the listening position back to either one of the amplifiers. The pair of amplifiers are connected by a second umbilical, ensuring that the absolute phase of both amps switches simultaneously when the remote switch is actuated.

The chassis and the potting cans for the transformers are fabricated from welded $\frac{3}{16}$" and $\frac{5}{16}$" aluminum plate and powder coated with a rich, chocolate brown colored, textured finish.

SC 280.2 Power Amplifier

Fig. 2 - Block Diagram

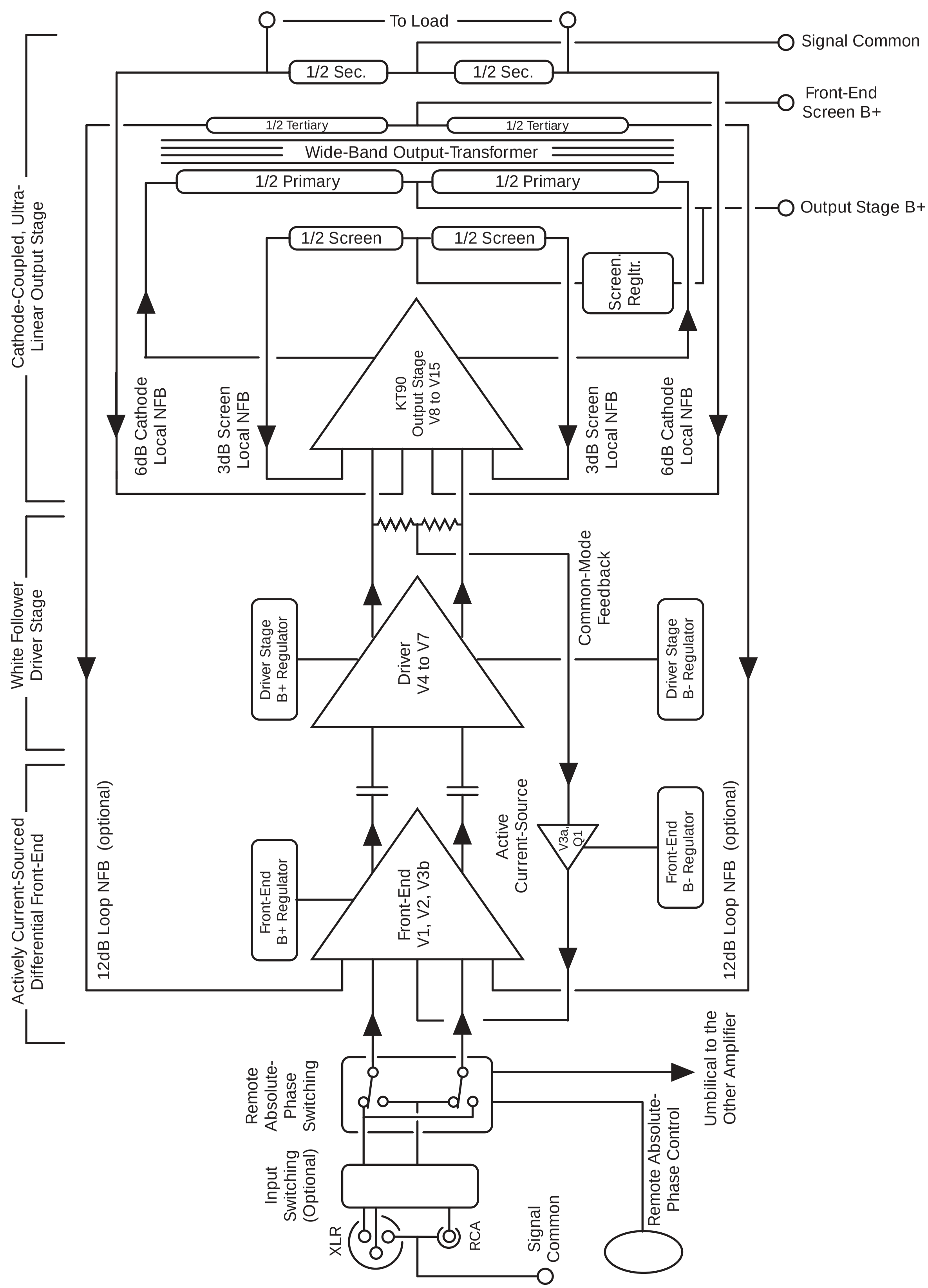
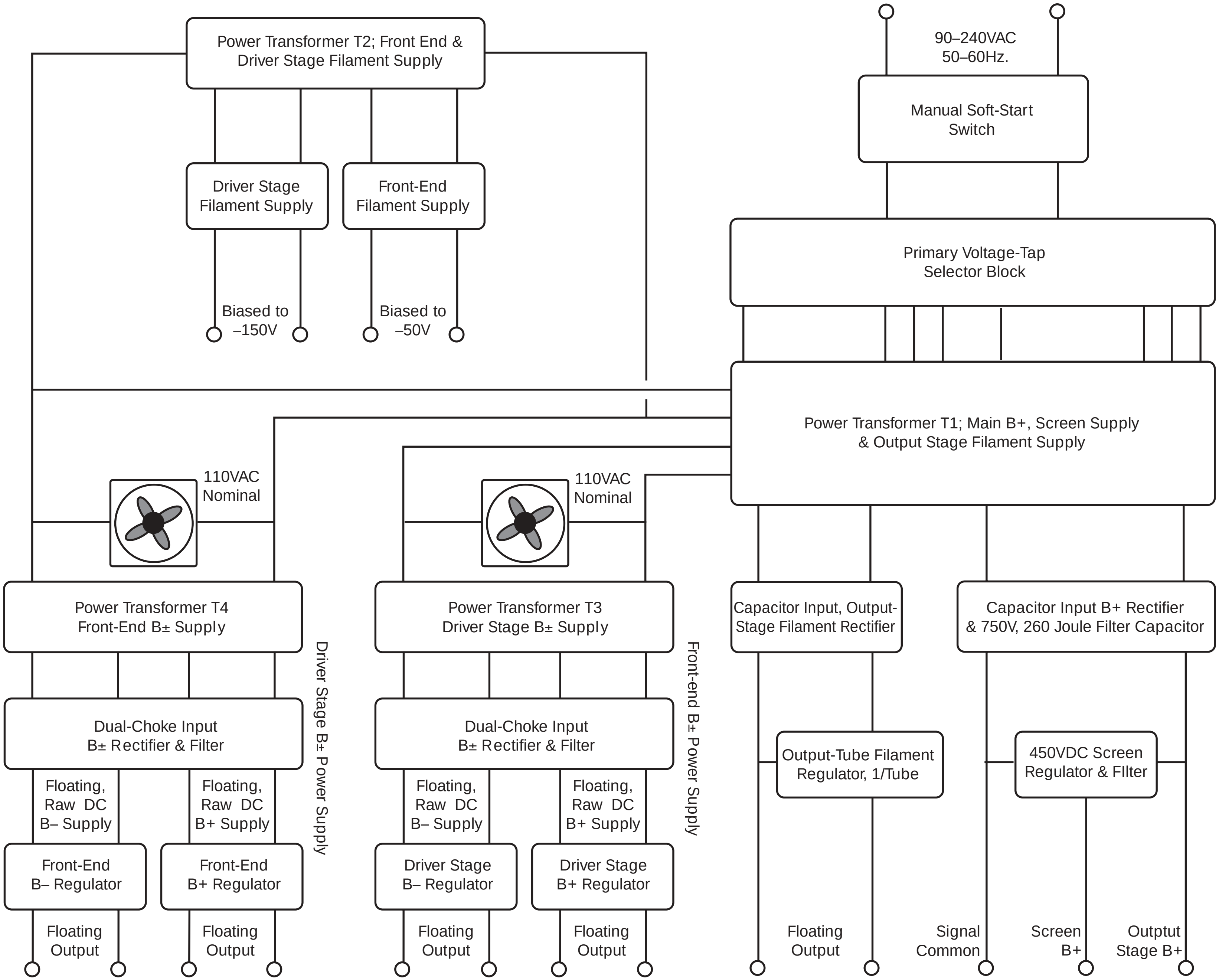
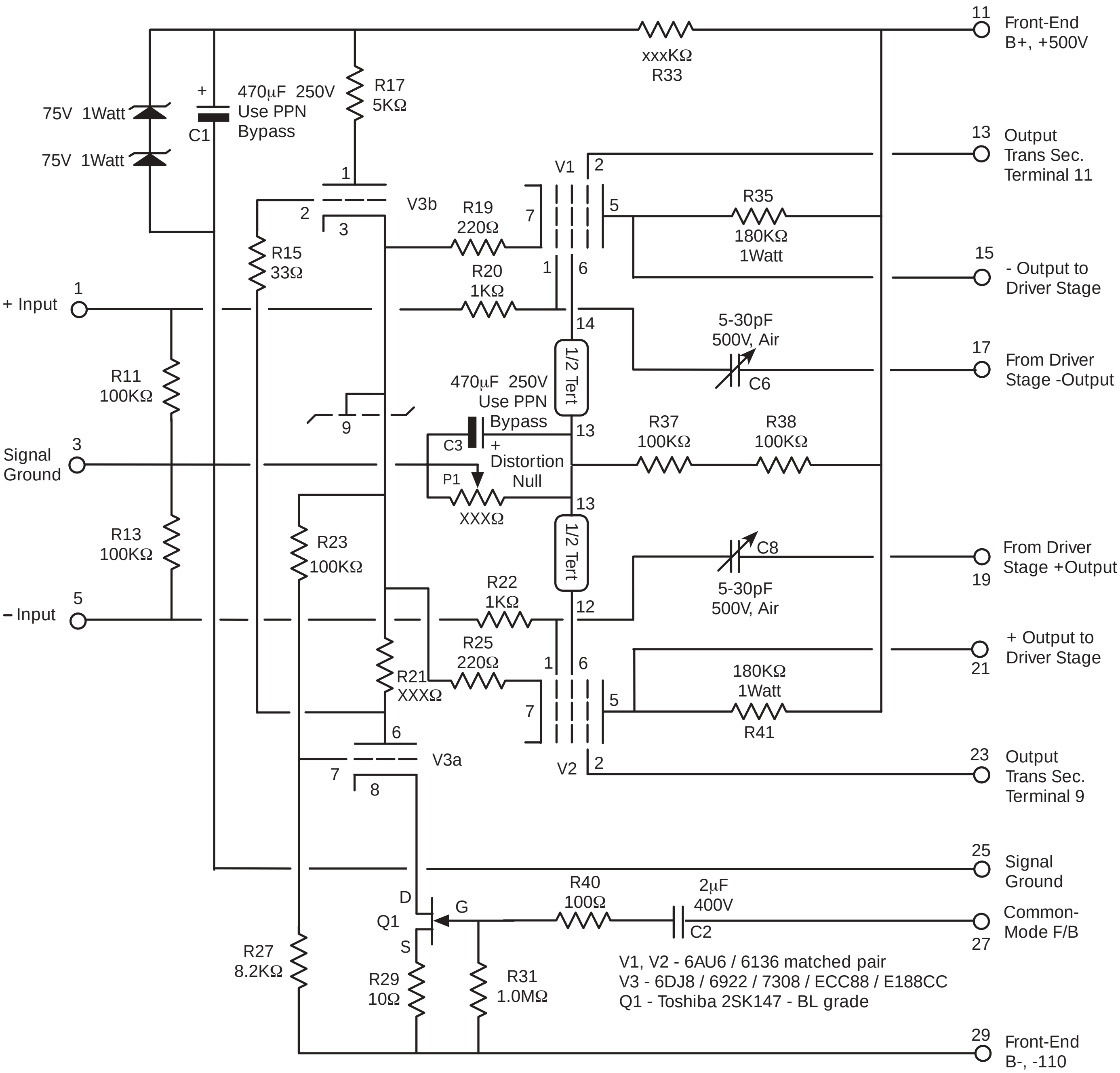


Fig. 3 - Power Supply Block Diagram



SC 280.2 Power Amplifier

Fig. 4 - 6AU6 Differential Front End - Component Values



Figs. 4 & 5. The front-end of the “loop-feedback version” of the SC 280.1 is shown in these two figures.

This stage is an actively current-sourced differential amplifier. The differential-pair consists of a pair of 6AU6 pentodes matched for both control-grid-to-screen-grid and screen-grid-to-plate transconductance. As pentodes matched this way are difficult to locate PEARL has undertaken the task of making these available to SC 280 builders.

An active current-source is formed by $V3_{a\,\&\,b}$ and Q_1. $V3_a$ and Q_1 form an ultra-linear cascode—see *Audio Note* 15.0 for more details—that provides about 50dB of gain. $V3_b$ acts as both an active plate load for $V3_a$ and as a low impedance driving-point for the common cathode connection of V_1 and V_2.

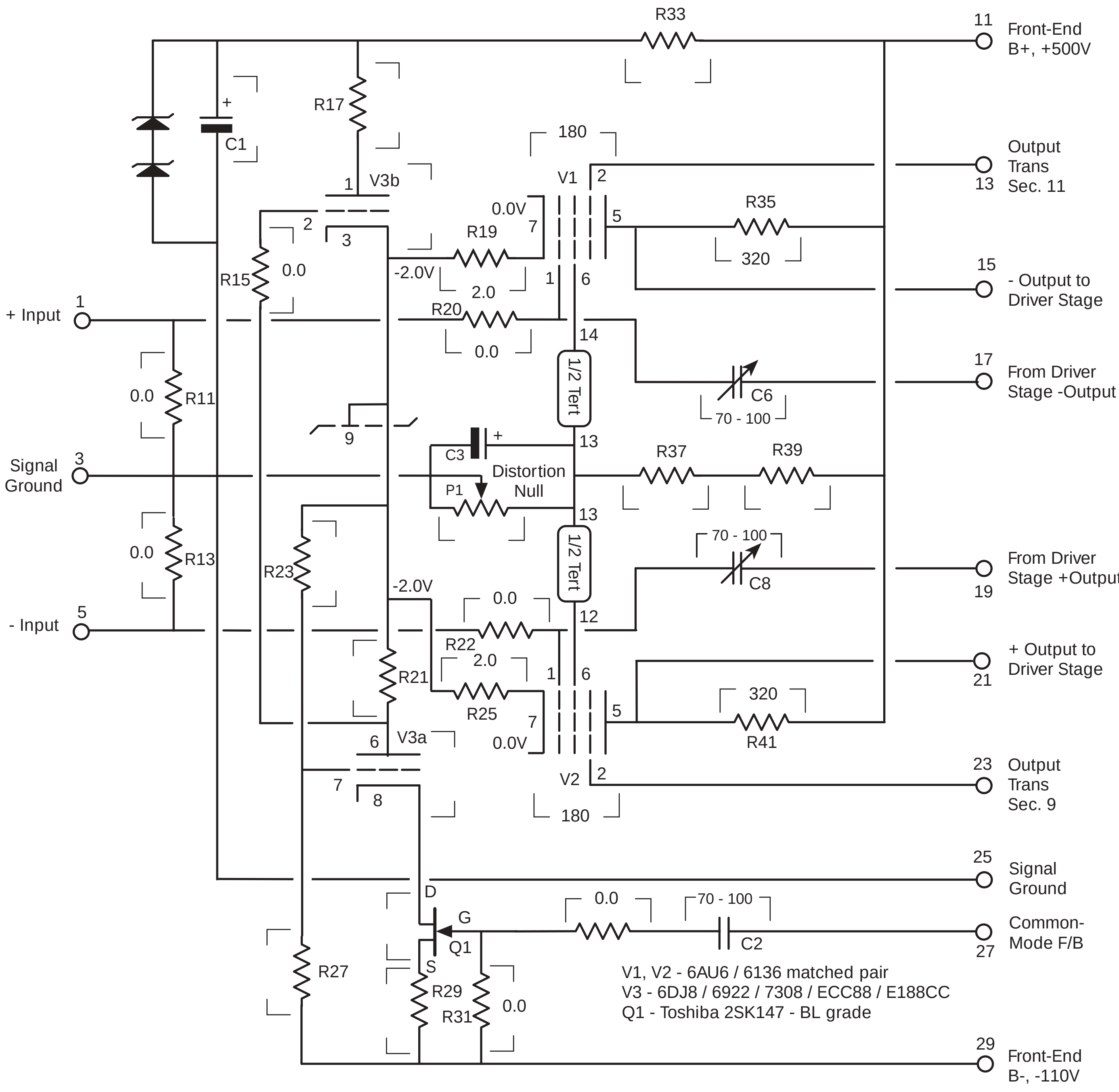
Approximately 12dB of voltage feedback is applied to the screen grids of the differential-pair from the tertiary feedback winding on the output transformer. This has an advantage over feedback more conventionally applied into the cathodes in that the reduced output impedance of the stage derived by the method shown has the effect of allowing the natural tendency of negative voltage-feedback to decrease output Z to work with itself. The higher output Z created by cathode NFB works against the Z-lowering effect of NFB.

The *Distortion Nulling* pot shown must be adjusted to place the optimum DC potential on the screens. A variation of only a few volts will yield major changes in the distortion figures shown by this highly linear circuit.

continues →

SC 280.2 Power Amplifier

Fig. 5 - 6AU6 Differential Front End - DC Voltages



The suppressor grids are connected to the secondary of the output transformer for the reasons outlined on page 6 of *Audio Note* 23.1, under the section heading “The Output Stage”.

The air-dielectric trimmer capacitors C6 & C8 provide high-frequency, negative voltage-feedback into the control grids of the differential input-pair.

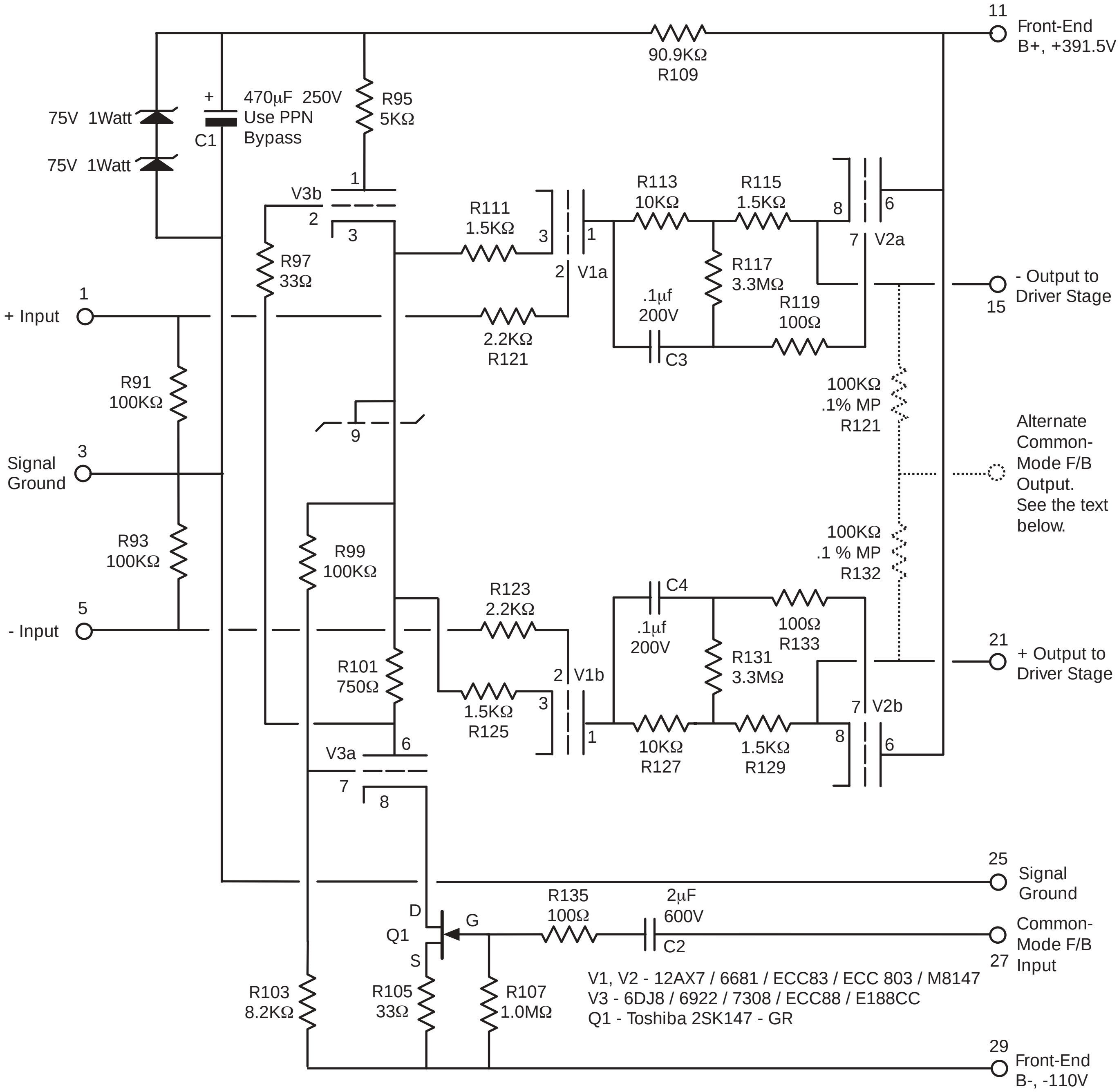
The loop feedback voltage applied to the screens from the tertiary winding of the output transformer appears there with a phase shift that increases with frequency and an amplitude that decreases. If the phase of the screen-feedback voltage ever swings far enough to become positive—rather than negative—feedback, instability of the entire amplifier can occur. To counteract this possibility, a subsidiary feedback voltage is taken from the outputs of the White-follower driver stage and applied to the control grids of the 6AU6's. This has the effect of “overpowering” any weak, regenerative feedback voltages that may appear on the screens.

At the high frequencies where both the potential problem of instability and its remedy, subsidiary feedback, are active, amplifier stability—rather than ultimate linearity—is the condition of interest. Consequently it is an acceptable practice to derive a feedback voltage from the stage immediately following the front-end even though doing so effectively drops the output stage out of the feedback loop.

Adjust the trimmers individually to achieve the best possible high frequency square wave response.

SC 280.2 Power Amplifier

Fig. 6 - 12AX7 Differential Front-End - Component Values



Figs. 6 & 7 The alternate "no loop feedback" version of the SC 280.1 front-end is shown in these two figures. As with the "loop feedback" version, the differential amplifier shown above uses an active current-source to create an accurately push-pull pair of signals to drive the output stage. These are created from an input signal that can be either single-ended or balanced.

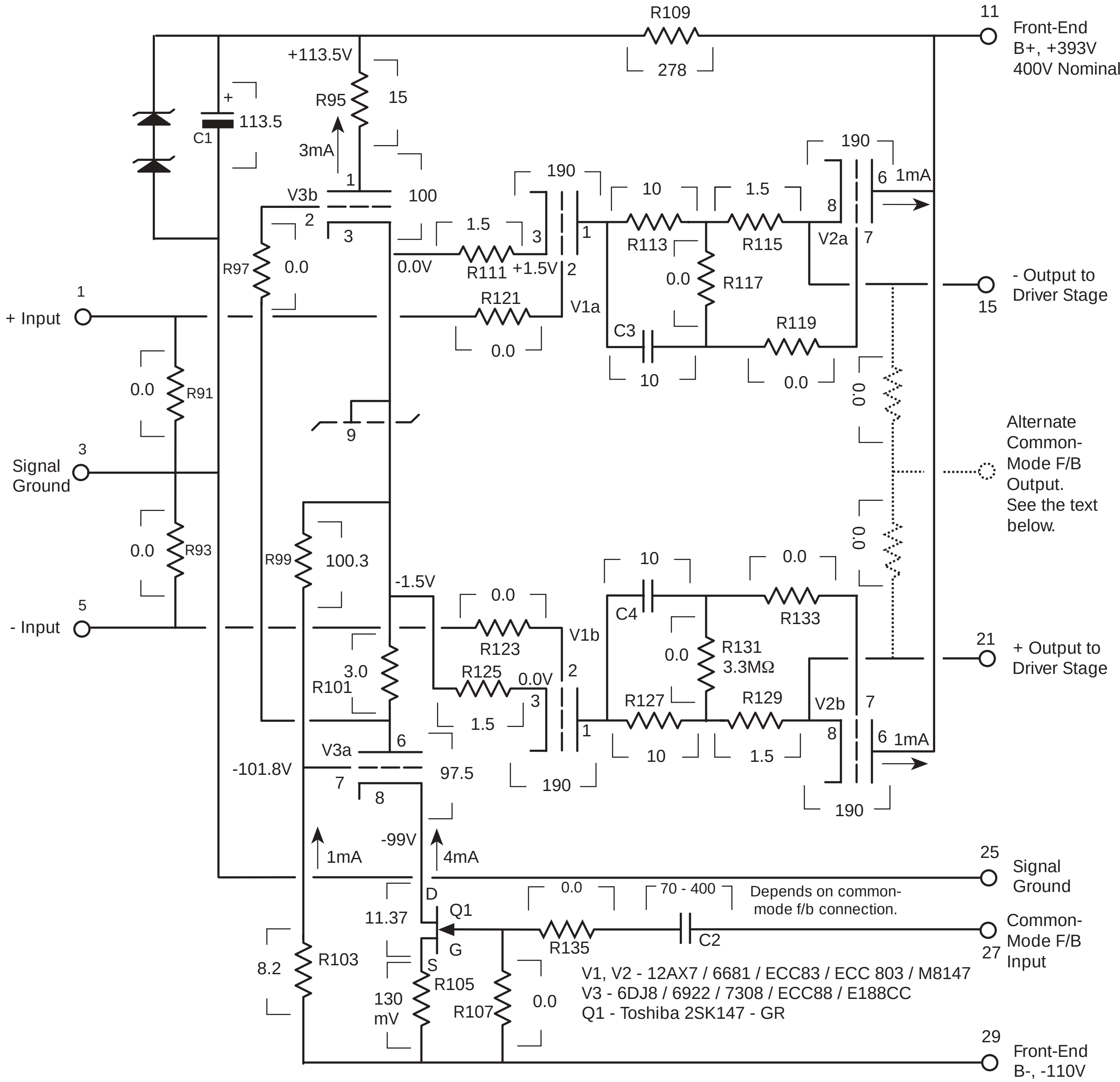
When driven from a balanced source, the possibility that the input signal may be contaminated by common-mode voltages arises. Such error voltages effect the push-pull input-pair equally and cause a shift in both voltages in the *same*—rather than the desired anti-phase—direction. The result—in the output of a differential amplifier with low common-mode rejection—is that the zero-volt position of the amplified waveform is moved away from zero. As previously stated, this has an undesirable effect on the vital information near the waveform's zero-crossing region. Low-level detail and ambiance data suffer most severely from common-mode contamination.

One of the actions of the active current-source is to ensure that any common-mode errors appearing at the output of the driver stage are substantially reduced in level. This effect maintains with imperfectly matched differential-pair devices and over time, as the tubes age. The typical common-mode rejection ratio of this circuit is 65 - 80dB from very low—sub 1Hz.—frequencies to well over 250kHz.

If you are opposed to feedback loops around multiple stages, the active current source can still be used. *continues →*

SC 280.2 Power Amplifier

Fig. 7 - 12AX7 Differential Front-End - DC Voltages



Disconnect the "Common-mode F/B Input" from the 100KΩ matched-resistor pair in the driver stage and connect it to the "Alternate Common-mode F/B Output" shown in dashed lines in the front-end schematic. Don't use driver stage outputs 17 and 19.

If you don't want to build the driver stage and prefer instead to work up your own biasing scheme for the output stage, the 12AX7 front end will drive an output stage using push-pull, parallel-connected EL34s. Be *certain* to use the pentode EL34 and not the tetrode 6CA7 if you build this way

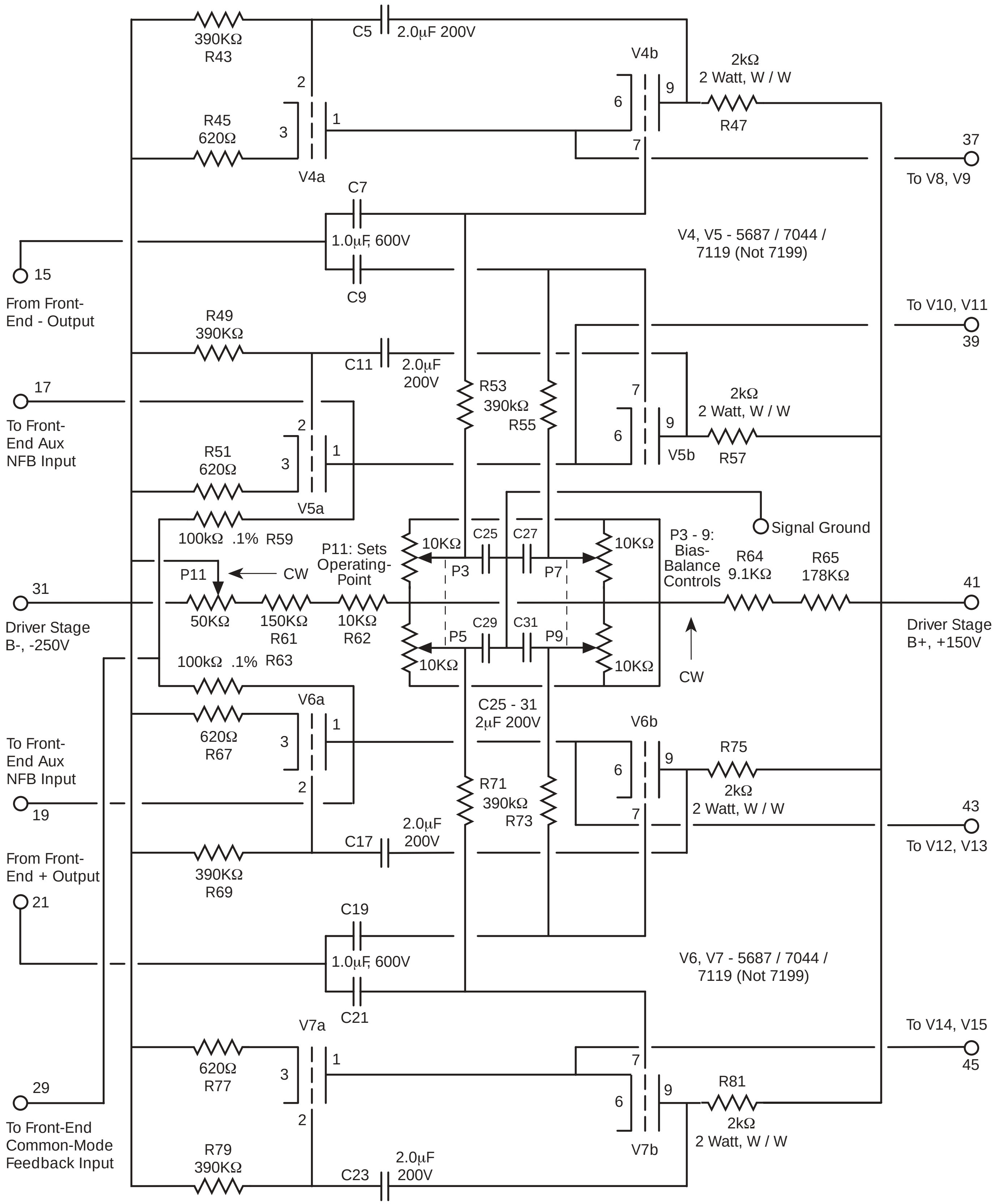
Worth repeating is the fact that even order distortion products show up as common-mode errors in most balanced circuitry. Here, they are greatly reduced in level by the active current-source.

The collection of R's and C's connected between the plates of V_1 and the cathodes of V_2 causes V_2 to act as a low output-impedance high-value plate load "resistor" for V_1. The use of a virtual high-value plate-load or so-called "current source" creates a number of benefits. The voltage gain of V_1 rises to 80, nearly the μ of a 12AX7, while the small-signal distortion drops, typically, by an order of magnitude. The output impedance drops from several hundred KΩ to 10KΩ, while the bandwidth of the circuit—into high impedance loads—is much improved, running out to well over 1MHz. with R121/3 shorted.

Please see *Audio Note* 14.0 *"The Active Plate Load; A New Level of Performance from the Dual Triode"* for a full explanation of the operation of this circuit.

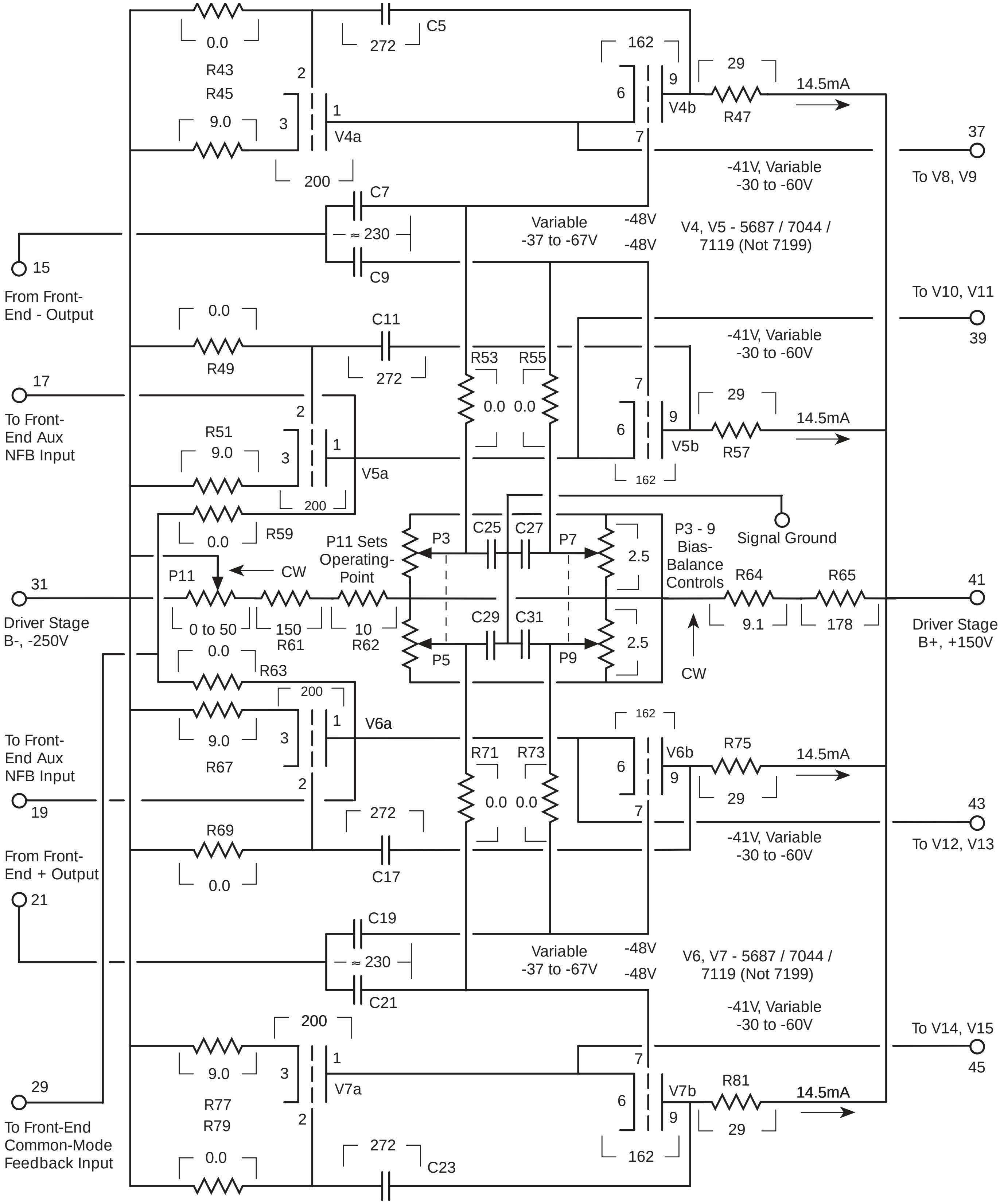
SC 280.2 Power Amplifier

Fig. 8 - White-Follower Driver Stage - Component Values



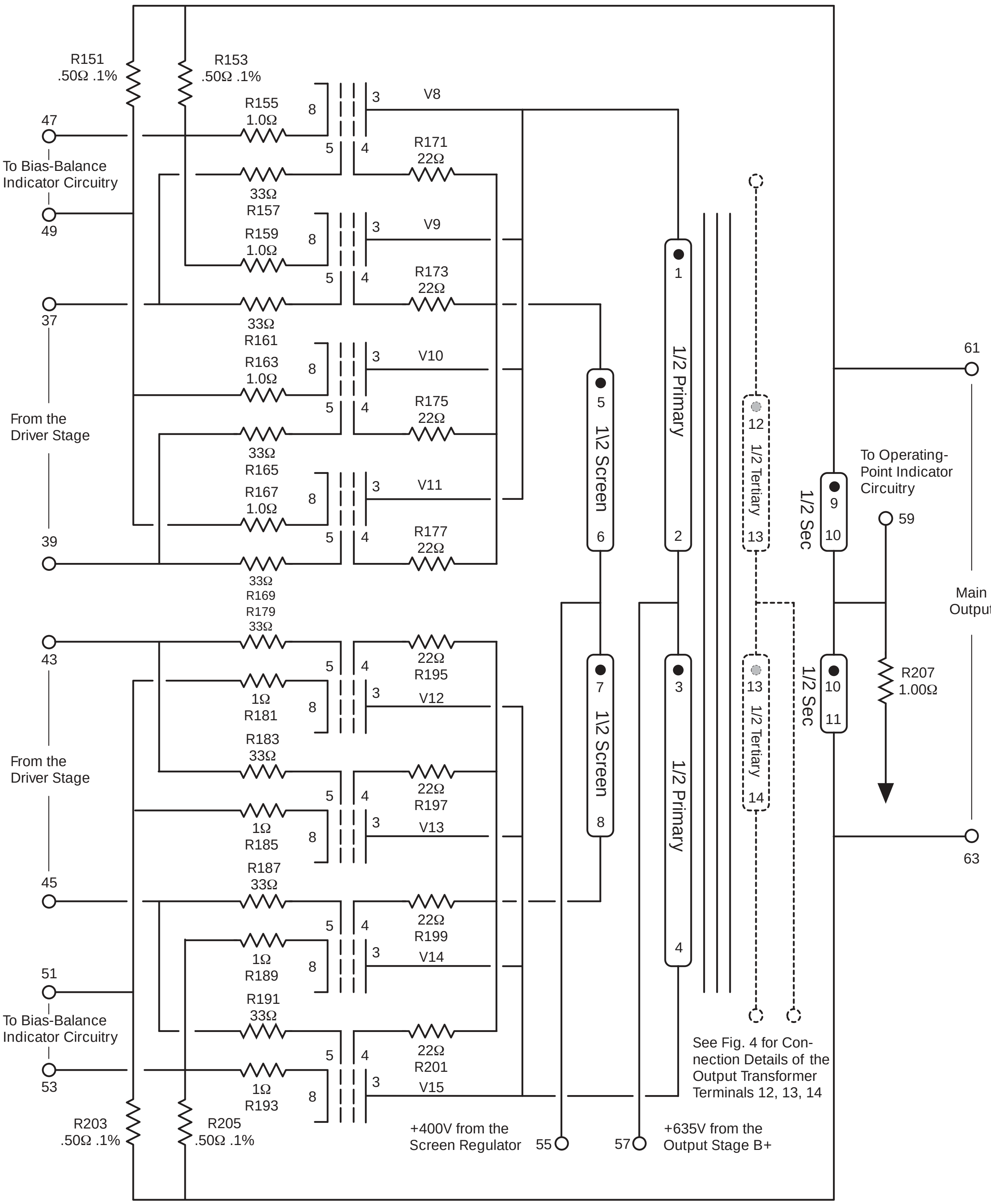
SC 280.2 Power Amplifier

Fig. 9 - White-Follower Driver Stage - DC Voltages



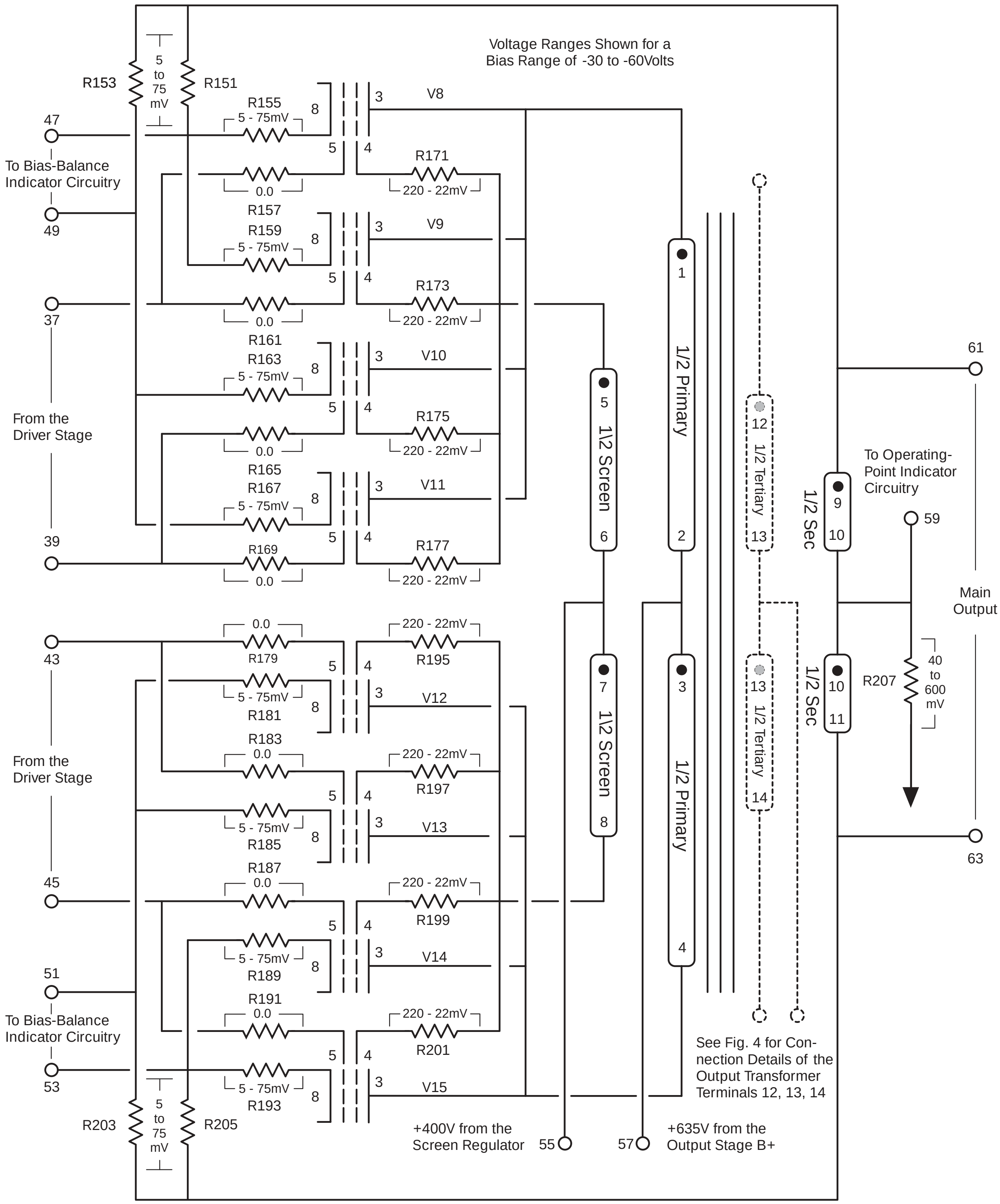
SC 280.2 Power Amplifier

Fig. 10 - KT90 Output Stage - Component Values



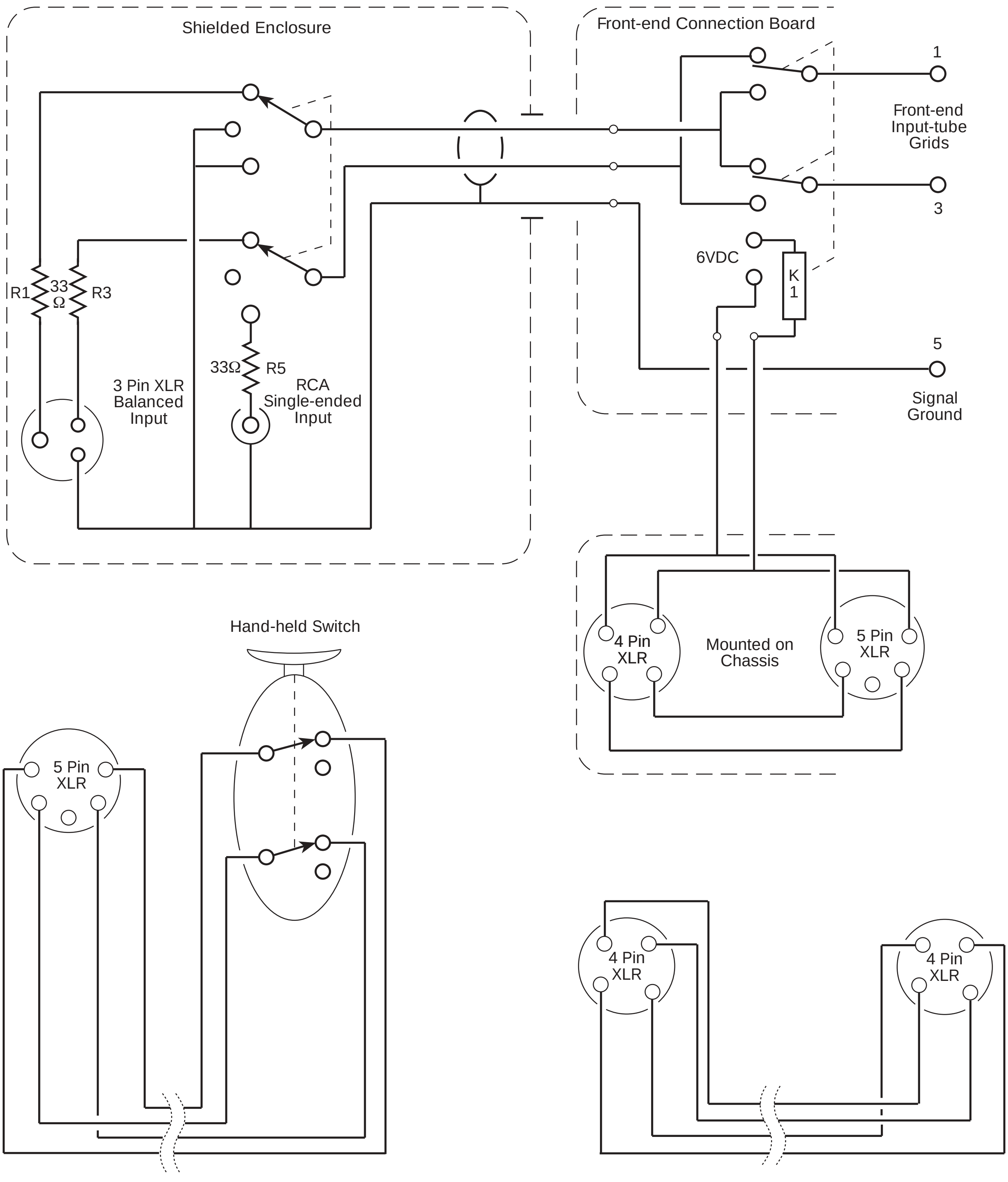
SC 280.2 Power Amplifier

Fig. 11 - KT90 Output Stage - DC Voltages



SC 280.2 Power Amplifier

Fig. 12 - Input Switching & Remote-Absolute-Phase Inverting Control



Absolute Phase Remote Control Umbilical

Amplifier Inter-connecting Umbilical

Part IV — This circuit, the "long-tailed–pair", is considered by the author to be the best of the phase splitters.

PUSH - PULL DRIVERS

BY GEORGE FLETCHER COOPER

A lightly edited, redrawn and re-typeset version of an article that appeared in Radio-Electronics *in March, 1953.*

IN THIS SERIES OF FOUR ARTICLES about the phase-splitting stage that provides the link between a single-ended voltage amplifier and a push-pull *[output]* stage, we have examined several poor circuits and two classes of acceptable circuits. We have seen that a single tube with a split load is good if you do not need too much drive while the anode-follower or *see-saw* circuit gives more output, but less gain. For conventional audio work, the tubes and supply voltages you plan to use will determine whether you take a single-tube or two-tube phase-splitting approach.

There are some special jobs, though, for which these push-pull driver types are not suitable. The most important of these is when you want to go down to extremely low frequencies, or even all the way to zero frequency. The anode follower (see Part III, in the February issue) includes one coupling capacitor, so that it will not stay balanced once the capacitor starts to take control; the split-load circuit (Part II, January, 1953) has the disadvantage that the two output terminals are at different DC potentials. A symmetrical direct-coupled deflection amplifier for a cathode-ray oscilloscope calls for a phase splitter that provides two outputs at the same average DC potential, because any unbalanced difference will pull the undeflected spot away from the center of the tube.

THE LONG-TAILED–PAIR

One phase-splitting circuit is particularly good for cathode-ray oscilloscope work. It has other uses too, but we shall come back to those after we have examined its characteristics. The circuit itself is known as the "long-tailed–pair" or more prosaically, as the Schmitt cathode-coupled phase inverter. If you look at the basic circuit in Fig. 1, you will see why the name "long-tailed–pair" was adopted. In a rather interesting way, this circuit is related to the anode-follower discussed last month. That circuit, you may remember, could be described simply by saying that the second tube is driven by the *difference* in the *plate-voltage* swings of the two tubes.

In the long-tailed–pair, the second tube has its grid grounded and the effective drive to this tube is applied to its cathode. The driving voltage is equal to the cathode resistance R_k multiplied by the difference in plate currents of the two tubes. Referring again to Fig. 1, suppose we raise the potential of grid

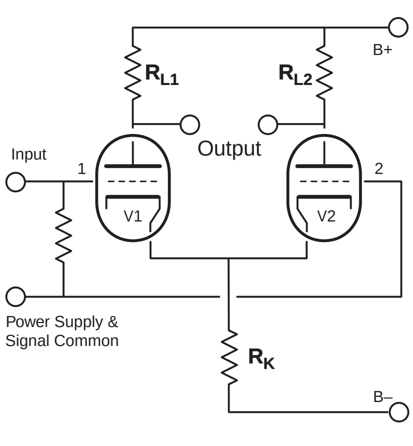


Fig. 1. Circuit of the "long-tailed–pair" cathode-coupled phase inverter.

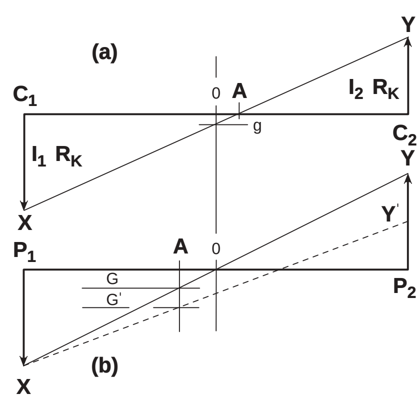


Fig. 2. (a) Current–see-saw diagram of the "long-tailed–pair". (b) See-saw diagram of balanced-voltage phase inverter reprinted from last month's article for comparison with the balanced-current type.

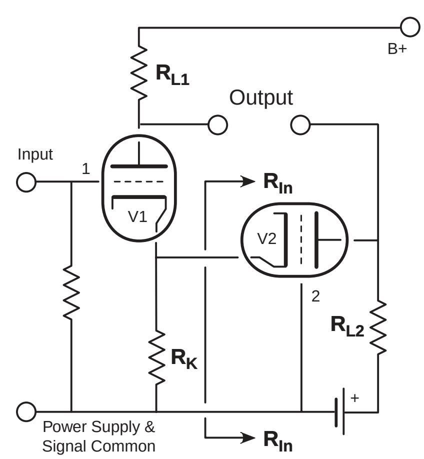


Fig. 3. The "long-tailed–pair" redrawn to show the grounded-grid operation of V2.

1 by 1 volt, thereby causing an extra current I_1, to flow through V_1-R_{L1}. The current in V_2-R_{L2} will *drop* by an amount I_2, and the cathode potential will change by $(I_1 - I_2)R_k$. This *change* is the input to V_2, and if R_k is very large, we can have $(I_1 - I_2)$ very small and still get the neccessary drive at V_2's cathode. In fact, we can draw Fig. 2a for comparison with Fig. 2b, which is the *see-saw–voltage* diagram from last month's article. Looking at Fig. 2a, you will notice that in order to get any drive at all to V_2, it is necessary to have unequal values of I_1 and I_2 (C_1-R_k and C_2-R_k)and the drive for V_2 is represented by the value O-g.

The only way to get equal swings at the two plates is to use slightly different plate-load resistances R_{L1} and R_{L2}. We can do some very simple calculations to see the sort of differences we can expect in the two values.

At plate 1, we have a voltage swing of I_1R_{L1}. The input to V_2's cathode is $(I_1 - I_2)R_K$, which gives a plate current of $g_m(I_1 - I_2)R_K R_{L2}$, which is, of course, also equal to I_1R_{L2}. Therefore, $g_m I_1 R_K R_{L1} = I_2(1 + g_m R_K)R_{1,2}$ and:

$$\frac{I_1}{I_2} = 1 + \frac{1}{g_m R_k}$$

Since we want:

$$I_1 R_{L1} = I_2 R_{L2}$$

we have:

$$\frac{R_{L2}}{R_{L1}} = \frac{I_1}{I_2} = 1 + \frac{1}{g_m + R_k}$$

Typical values for this circuit are:

R_k = 5KΩ and
g_m = 5ma/v (5,000 μmhos),

giving a ratio: R_{L2}/R_{L1}=1.04.

This means that, even without unequal load resistors, we get a balance correct to 4%, which is quite good.

The Rigorous Approach

This result, obtained by very simple reasoning, applies quite well if V_1 and V_2 are pentodes, though it leaves a rather tricky gap if you wonder how to bypass the screen grids right down to zero frequency. It shows that the circuit looks good, anyway. And now we come up against the main problem of the writer of technical articles: do you want it easy, or do you want it right? Sometimes the writer can put in all the mathematics—if he is French he seems to put in nothing but the mathematics—sometimes he collects a dense mass of small-print equations at the end. Usually he must tack between the Scylla of editorial condemnation and the Charybdis of long-haired readers who write and point out the smallest deviation from rigour.

My preference is to start off with the exact mathematical solution and then simplify. That way, any reader who wants to get more detail can build up on

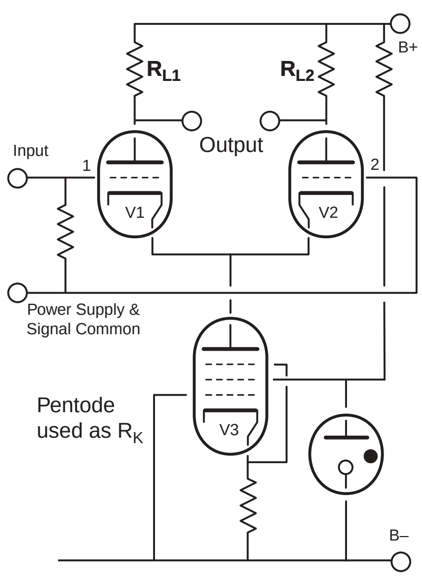


Fig. 4. A high-impedance pentode as a common–cathode-resistor swamps out variations in tubes 1 and 2 that might unbalance the phase-inverter currents.

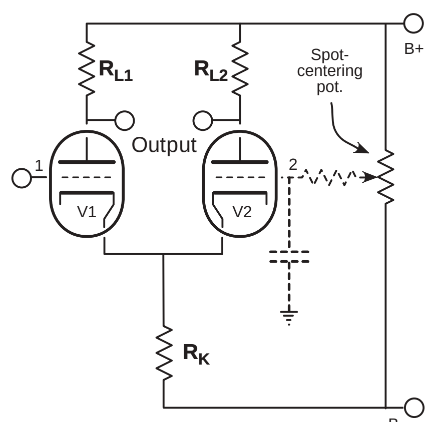


Fig. 5. When using the "long-tailed–pair" to feed the balanced deflecting plates of a cathode-ray oscilloscope, the grid of V2 may be returned to an adjustable, DC divider to restore the beam to the center of the 'scope screen.

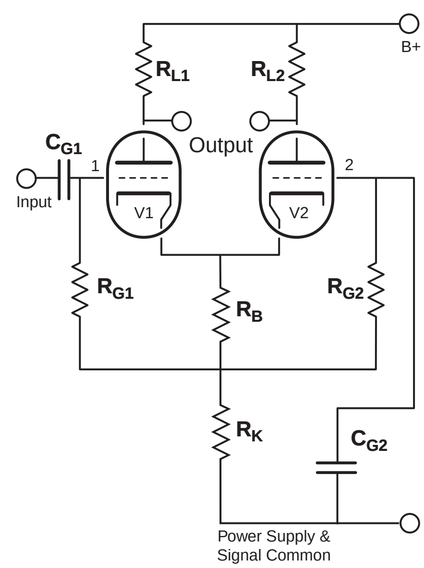


Fig. 6. One method of eliminating the negative voltage supply for the cathode return. R_B is tapped off the total cathode resistance at the desired bias point.

the sound foundations I have provided. If I simplify *first*, there is nothing to build on later.

It will be much easier to understand the long-tailed–pair if we redraw the circuit in the form of Fig. 3. This is not as elegant and symmetrical as Fig. 1 but it is a good deal more useful. V_2, you now see, is a grounded-grid stage, a subject we discussed recently in RADIO-ELECTRONICS (October, 1952 issue). I shall save myself a lot of trouble by using some of the results I worked out in that article. In particular, the input impedance (R_{In}), looking in at the cathode, is $R_{L2} + R_{p2} / 1 + \mu_2$, and the gain is $(1 + \mu_2)\, R_{L2} / R_{L2} + R_{p2}$.

First, let us look at the stage V_1. The tube has a plate load R_{L1}, and there is some negative feedback because of the unbypassed cathode resistor. If we represent the cathode circuit (R_K in parallel with the input resistance of V_2) as R_x the gain is just $\mu_1 R_{L1} [\, R_{L1} + R_{p2} + (1 + \mu_1)\, R_x \,]$ and the voltage across the cathode circuit is R_x / R_{L1}, times the voltage at the plate. So if we have 1 volt at the grid of V_1 we have $\mu_1 R_{L1} / [\, R_{L1} + R_{p1} + (1 + \mu_1)\, R_x \,]$ volts at plate l, $\mu_1 R_x / [\, R_{L1} + R_p + (1 + \mu_1)\, R_x \,]$ volts at the common cathode, which we can write as V_x, and $[\,(1 + \mu_2) R_{L2} / + R_{p2} \,]\, V_x$ at plate 2.

We want the voltage at plate 1 to be the same as the voltage at plate 2, so we put:

$$\frac{\mu R_{L1}}{R_{L1} + R_{p1} + \left(\mu_1 + 1\right) R_x} = \left(\frac{\left(1 + \mu_2\right) R_{L2}}{R_{L2} + R_{P2}} \right) V_x$$

and:

$$V_x = \mu_1 R_x / R_{L1} + R_{p1} + (1 + \mu_1) R_x$$

There is one more equation, because R_x is made up of R_k in parallel with R_{in}—$[\,(R_{L2} + R_{p2}) / (1 + \mu_2)\,]$— so $R_x = R_K (R_{L2} + R_{p2}) / (R_{L2} + R_{p2} + \mu R_k)$.

That last paragraph is difficult but true. If you assume the tubes are similar and struggle with it for a while, you'll come out with the equation for true push-pull balance:

$$R_{L1} / R_{L2} = (1 + \mu)\, R_K / R_{L2} + R_{p2} + \mu R_k$$

This doesn't look too bad, and if we assume further that we have pentodes, so that μ and R_{p2} are very large $1 + \mu \approx \mu$ and $g_m = \mu / R_p$, we arrive at:

$$R_{L1} / R_{L2} = g_m R_K / (1 + g_m)$$

which is equivalent to the answer given by our very simple treatment.

It's interesting to compare the exact solution to the simple one for a typical case: consider a 12AT7 with R_p = 10KΩ, μ = 50, g_m = 5,000 μmhos and R_K = 5KΩ. We can choose R_{L2} = 50KΩ and we obtain:

R_{L1} / R_{L2} = 51 x 5,000/50,000 + 10,000 (50 x 5,000)
= 255,000/310,000 = .823 whereas with the simple formula we had: R_{L1} / R_{L2} = 1/1.04 = 0.96.

There is quite a difference between a 17.7% and a 4% unbalance; we are obviously justified in adopting the more rigorous approach. Energetic readers may care to consider what happens if V_2 ages and R_{p2} gets bigger: as R_{p2} represents only 10,000 in the total 310,000 of the numerator it cannot be a very serious factor. If μ_2 gets smaller, things are rather more complicated, because μ_2, appears in both denominator and numerator; anyway, μ_2 is more nearly constant during the life of the tube.

OSCILLOSCOPE APPLICATIONS

For applications in which a good, stable balance is needed, a much higher value of R_K would be used. As an example, we might take R_K = 50KΩ, when:

$$R_{L1} / R_{L2} = 2{,}550{,}000 / 2{,}560{,}000 = 0.996$$

and the unbalance is only about .5%. Changes in tube characteristics can only unbalance the circuit to the same limited extent assuming that we start with correctly proportioned load resistors, so that we have a very satisfactory circuit here but 50KΩ is a *very* long tail.

The reader who has been watching these numbers carefully may be getting a little worried. If each tube draws 5mA, there will be a total of 10mA through the cathode resistor. all this talk of 50KΩ implies a drop of 500V across R_K. Even if we drop R_K to 20KΩ, we still need a –200V supply if we are to work with the grids around ground potential. This is not a serious matter in oscilloscope circuits, because negative high-voltage supplies are generally used. In some other circuits, where the grid of V_1 is connected directly to the plate of a preceding tube it is an advantage to have the whole tube circuit lifted up above ground. But 200–500V is rather high and when an extra-large cathode impedance is needed for special high-balance jobs, special circuit tricks are usually adopted.

One very important circuit uses a pentode in place of R_K. For example, a 6AQ5 with 200V on the screen grid will pass 10mA with only 20 volts on the plate, but the impedance so far as constancy of current is concerned will be very high, certainly above 100KΩ. This means that we can make $R_{L1} = R_{L2}$, and still have a virtually perfect balance. The form the circuit takes is shown in Fig. 4, which also shows a voltage regulator tube as a screen bypass, to ensure operation down to zero frequency.

Another way of providing the necessary high impedance in the cathode circuit is to use a saturated diode for R_K. The tungsten-filament diodes used as noise sources in receiver testing give full emission (saturate) at a relatively low plate voltage, and increases in plate voltage give almost no change in current. The only disadvantage in using them is that the diode current depends on the filament temperature, so the filament current must be stabilized.

Our purpose in seeking such highly accurate balance is not to get 10.01 watts from an amplifier instead of 10 watts, but to meet the requirements of some special measuring instruments. If you look back at Fig. 1, you will see that apart from the ground connection of the grid of V_2, the circuit is absolutely symmetrical. In fact, this ground connection is there only because we assumed a grounded input: we are really using the voltage $G_1 – G_2$ as the input, and deriving two, equal, antiphase outputs from this. Suppose, however, that we connect both grids together, and then apply a signal. There will be an in-phase *[common-mode–error]* signal at the plates, but it will be relatively small, since the large cathode resistance R_K provides a great deal of feedback. We can easily calculate what will happen, because we can assume that each tube has $2R_K$ in its cathode and then treat one tube alone. From grid to plate, the gain is:

$$\mu R_L / [R_L + R_P + (\mu + 1)\, 2R_K].$$

Connecting a 12AT7 (which has a μ = 50 and an R_p = 10KΩ), in circuit with an R_L = 50KΩ and an R_K = 50KΩ gives a gain of about 0.5. So for push/push *[common-mode]* input the gain from grid to plate is less than unity. For push-pull *[differential-mode]* input, applied *between* the grids instead of to grids in parallel, the gain is 30, so the balance *[common-mode–rejection]* ratio is 0.5:30, 1 to 60 or –35.6dB. If a pentode is used for the cathode resistor this figure can be increased still more.

This circuit is used in electroencephalography: the two grids are connected to electrodes applied to the head of a patient and the tiny brain-currents produce a push-pull *[differential-mode]* voltage between the electrodes. Stray 60Hz. fields produce a relatively large push-push *[common-mode]* voltage that must be eliminated,because it would mask the brain signals even if it did not overload the final stages of the recording amplifier.

For use with oscilloscopes there are two possibilities. In the first, the two grids can be regarded as the two input terminals, and we have the feature that push-push voltages are discriminated against, while push-pull voltages are applied to the deflecting plates of the oscilloscope tube. This is excellent if you wish to work around zero voltage. But if you are interested in the variations of a voltage which is always well away from zero, this push-pull input is not very satisfactory, because the DC component of the input will deflect the spot away from the center of the screen. The arrangement of Fig. 6 is then more useful. The grid of V_2 is connected to point A, which provides a positioning voltage to bring the spot near the center of the screen. The grid of V_1 takes the input signal, which is converted to push-pull to avoid defocusing and trapezium distortion *[key-stone-ing]*. A capacitor (shown in dotted lines) can be added to keep supply-hum off the grid of V_2, where it would be amplified as an ordinary signal A resistor (also shown dotted) is sometimes added to improve the smoothing. (In some oscilloscopes the time constant of this RC filter is so long that the spot goes on drifting long after you have taken your hand off the positioning control.)

Stages of this kind can be connected in cascade if you have generous power supplies. The subject is slightly outside our present field, but you can see that if the first pair is to operate at about zero grid-volts, the plates will be up at about +l00V, while the cathode resistor is returned to, say, –150V. The second stage grids are then at +100, so we have a 250V drop in the second cathode resistor (assuming this also goes back to –150V. The second stage plates will be at +200–250 volts, so that the supplies needed will be +400V, 0 and –150V. You just apply Ohm's law to find the resistance values.

Although this circuit is not used much for audio amplifiers, we should examine how it *can* be used, and, in particular, how we can get away from this negative-voltage line. A simple and apparently symmetrical form of circuit is shown in Fig. 6. Bias is provided by R_B, which has its usual value for the tubes (about 100–500Ω) according to the operating conditions, and the grids return to the bottom of R_B through R_{G1}, R_{G2}. The coupling resistor R_K lifts the whole group of resistors up to perhaps +100V, so that blocking capacitors C_{G1}, C_{G2} are needed to connect the input and to provide the AC ground on grid 2. This circuit is not as symmetrical as it looks, as you will see if you consider it as redrawn in Fig. 7. At high frequencies, the grid of V_2 is grounded but at low frequencies, when C_{G2} is no longer a low impedance, the grid is returned to somewhere between ground and the top of R_K. At zero frequency, the coupling resistance is down to R_B, which is too small to provide any satisfactory sort of balance. In practice this means that we must make $2\pi f C_{G2} R_{G2} >> 1$ at the lowest frequency we intend to use. We also need $2\pi f C_{G1} R_{G1} > 1$ if we are to get the signal into the circuit at all. At high frequencies the only sources of trouble are tube capacitance, in particular the grid-cathode capacitance which is effectively in parallel with R_K, since grid 2 is grounded.

If you now look back at Fig. 1 you will, no doubt, admire the elegant simplicity of the circuit: just three resistors, a pair of tubes, and that little *–Ve* sign. Every few years I come back to this point of indecision. It's a good simple circuit but where will I get that negative supply? For special jobs, with double input, it is possible to elaborate the long-tailed–pair to give a really well-balanced system, though the balance is usually not as good as you can get with a transformer. But the long-tailed–pair stays balanced down to zero frequency.

I hope that in these four articles I have succeeded in making it clear that it doesn't cost any more to make your push-pull circuit really balanced. If you want to use a reasonable amount of negative feedback, the shoddy circuits described in Part I will add to your troubles and, as far as I can see they, amount to nothing more than a public avowal that you "couldn't care less." If an amplifier is designed on that basis, I would expect it to be pretty badly constructed as well—I'd stay well away from it.

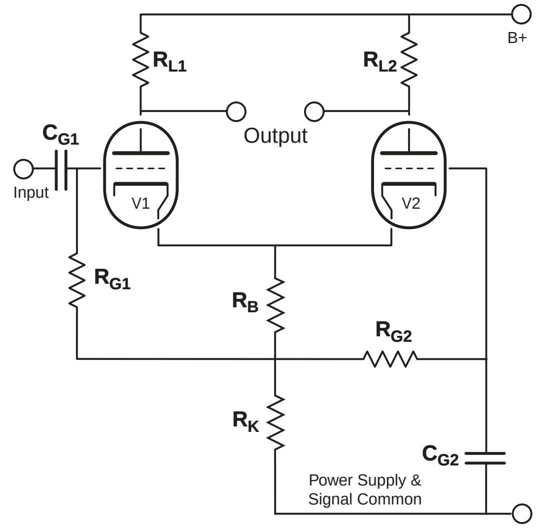


Fig. 7. One method of eliminating the negative-voltage supply for the cathode return. R_B is tapped off the total cathode resistance $(R_B + R_K)$ at the desired bias point.

Balanced Amplifiers of Highly Stable and Accurate Balance

A redrawn and slightly edited version of an original published by E.M.I. Laboratories

A KNOWN AMPLIFIER capable of developing a stable, balanced output free, to a considerable degree, from unbalanced components despite conditions of unbalance and possibly changing conditions of unbalance in its input circuit, employs a pair of amplifier valves having in their cathode circuit a common impedance. This impedance is normally a pure resistance connected at one end to the negative terminal of high tension supply and at the other directly to each cathode. Input signals can be applied to each of the control electrodes *[the control grids]* of the two valves and the control exerted upon the two valves is then determined by the signal variations present at the control electrodes and by such potential variations as may be developed at the connected cathodes. The latter variations depend upon the current variations in the two valves and upon the magnitude of the resistance forming the common cathode circuit. If the variations of cathode current are equal and opposite so that balanced currents are set up in the valves, then it is evident that the potential developed at the cathodes of the valves remains constant. If, however, the valve currents tend for any reason to depart from a condition of balance, the difference of these currents flowing in the common cathode resistance results in variations of the potential at the common cathode-connection, thereby setting up a differential control of the valves that tends to equalize their currents. An amplifier of this kind will produce a satisfactorily balanced output even when only one of its pair of control grids is excited. This requires a value of common cathode-resistance equal at least to the reciprocal of the mutual conductance of the valves and, with a resistance of such magnitude, the essential characteristic of the amplifier is its notable insensitivity to so-called *push-push*[‡] components of variation present at its input terminals.

It will be clear from these remarks that the desirable properties of the circuit depend upon the use of a common cathode resistance of large magnitude. It will also be appreciated that the use of such a resistance can make demands upon the source of high tension that are at least inconvenient, even if the waste of power can be tolerated. This is so, particularly if the valves take large currents, for then the inevitable drop of potential in the cathode resistance becomes very large.

It has accordingly been proposed to replace the common cathode-resistance by a valve, and not merely to do this, but also to control this valve from a point on a potential divider connected between the output circuits of the pair of valves. Thus, with similar valves and equal output resistances connected in their anode circuits, the potential divider may be connected directly between the anodes and the control taken from the midpoint of the divider.

With such an arrangement the stabilising effect of potential variations set up at the cathodes and resulting from an unbalanced condition in the output of the amplifier can be increased to a degree vastly in excess of that obtainable from an arrangement using simple a resistance for the common cathode circuit and not incurring a greater drop of potential in the cathode circuit. The amplifier may be driven by either a balanced input signal or by an unbalanced signal, in which case one of the two control electrodes of the differential-pair is effectively grounded. Fig. 1. illustrates a circuit for balanced-input operation and AC conditions. Fig. 2. shows a similar circuit arranged for DC operation.

The improved form of amplifier has one advantage in particular over the normal form in which the cathodes are coupled only by a passive impedance. This advantage is the ability to compensate automatically for load-circuit changes that would otherwise upset the balance of the output. For example, a reduction in load

‡common-mode signals, in today's parlance.

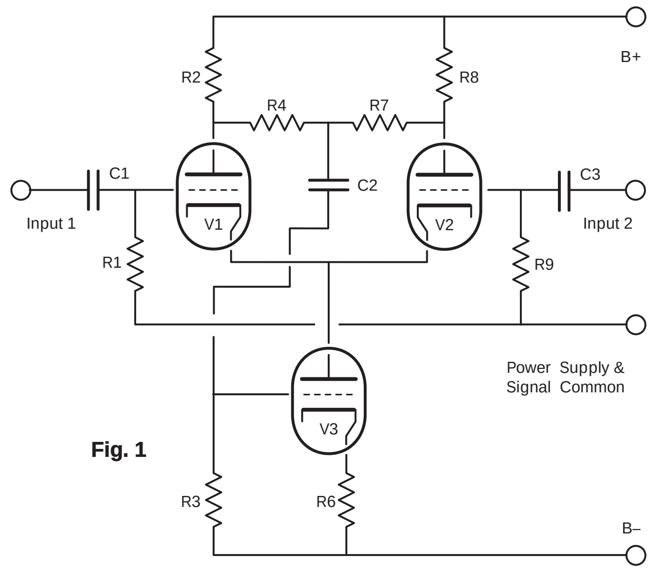


Fig. 1

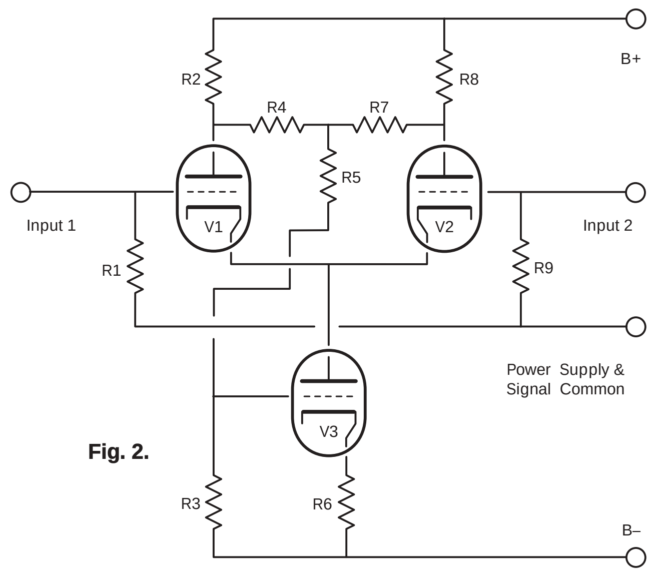


Fig. 2.

Figs. 1 & 2. V1 & V2 form the differential-pair while V3 drives the common–cathode-connection in such amplitude and phase as to force the voltage at the junction of R4 & R7 to nearly zero under widely varying conditions of load and input drive.

on one the anodes of the differential-pair effectively reduces the value of its anode resistance and accordingly, the signal-voltage generated across it.

If desired, the amplifier can be configured to deliver its output in a predetermined unbalance-ratio. In such an instance, the tap on the potential divider is appropriately displaced from the mid-point so that when the output is exactly in the required ratio no control is exerted on the common cathode-valve. In the figures it will be noticed that the common cathode-valve is provided with a cathode-load resistance. This is not merely to provide a convenient bias for the valve, but also to render its operation less dependent on changes in its characteristics and is simply a part of well known technique.

—Communication from E.M.I. Laboratories.

An Improved Type of Differential Amplifier

By J. C. S. Richards[†], B.Sc., Ph.D.

A lightly edited, redrawn and re-typeset version of an article that appeared in *ELECTRONIC ENGINEERING* in July, 1956.

A differential amplifier-stage capable of giving a high rejection ratio with unselected valves and components and without a balance control is analyzed,and a particular amplifier is described in some detail. The stage is particularly suitable for converting balanced to unbalanced signals.

CONSIDERABLE ATTENTION has been given to the design of differential amplifiers—that is, amplifiers having two ungrounded input terminals and which give zero or negligible output when like signals are applied between each of the input terminals and earth.

The ability of a differential amplifier to discriminate between like *[common-mode]* and unlike *[differential-mode]* input signals is expressed in terms of its *[common-mode]* rejection ratio, defined more precisely below. With most designs of differential amplifier, it is necessary either to use specially selected valves and components, or to have some form of balancing control to obtain high values of rejection ratio.

It is shown now that by means of a feedback circuit, it is possible to make a differential amplifier having an inherently high rejection ratio without the use of selected valves or a balance control. The addition of a single balance control makes it possible to attain an effectively infinite rejection ratio. The high rejection-ratio is maintained over a wide band of frequencies. The output of the amplifier can be taken push-pull or single-ended.

PERFORMANCE PARAMETERS

If a differential amplifier has a push-pull output, its performance can be described in terms of the differential gain M_d, the in-phase *[common-mode]* gain M_i, and the inversion gain M_v. [1, 2] If the input signals are e_1 and e_1', and the output signals are e_2 and e_2' (all measured with respect to earth), these parameters are defined as follows:

$$M_d = (e_2 - e_2')/(e_1 - e_1') \text{ for } e_1' = -e_1$$

$$M_i = (e_2 + e_2')/(e_1 + e_1') \text{ for } e_1' = e_1$$

$$M_v = (e_2 - e_2')/(e_1 - e_1') \text{ for } e_1' = e_1$$

If a push-pull output is used, the rejection ratio is M_d/M_v and the value of M_i is comparatively unimportant. However, if the early stages of the amplifier have a small value of M_i, the design of later stages is greatly simplified. In addition, the simplest way to obtain a single-ended output is to design a push-pull amplifier for which the overall value of M_i is very small, so that the differential properties of the amplifier are retained if the output signal is taken between one of the output terminals and earth.

STANDARD TECHNIQUES

The conventional type of differential amplifier stage is shown in Fig. 1. For this stage we have:

$$M_d = \frac{uR}{R + r_a} \quad \ldots\ldots(1)$$

$$M_i = \frac{\frac{1}{2}(\mu_1 - \mu_2)(R_1 - R_2)R_3 + \mu R(R + r_a)}{(R + r_a)\left[R + r_a + 2(\mu + 1)R_3\right]} \quad \ldots\ldots(2)$$

$$M_v = \frac{r_a^{\,2}(g_{m1}R_1 - g_{m2}R_2) + (\mu_1 - \mu_2)R(R + 2R_3)}{(R + r_a)\left[R + r_a + 2(\mu + 1)R_3\right]} \quad .(3)$$

These expressions are slightly simplified versions of the exact formula. R, r_a and μ are the mean values of $R_1 + R_2$, ra_1 and ra_2, and μ_1 and μ_2 respectively. They have been used where the exact values are not critical.

When R_3 is made large, both M_v and M_i become

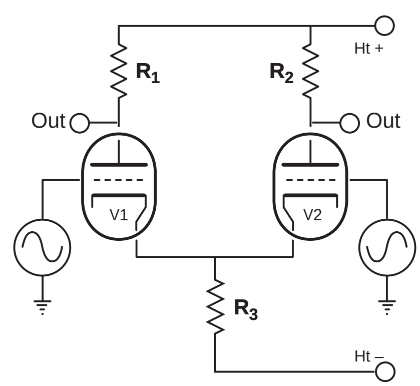


Fig. 1. A simple differential amplifier-stage.

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small although neither tends to zero. Now, for the whole amplifier, the value of M_i can be made very small by cascading several stages. On the other hand, M_v must be made small for each stage where the in-phase input signal is appreciable (in practice generally the first stage only) since large value for M_v means that an appreciable anti-phase signal is passed on to later stages. The main problem, therefore, is to make M_d/M_v for the first stage sufficiently large.

M_v depends on the balance between the parameters of V_1 and V_2 and between R_1 and R_2 but if R_3 is made very large M_d/M_v becomes approximately $(\mu + 1)\,\mu/(\mu_1 - \mu_2)$, which is large if μ is large, even when $(\mu_1 - \mu_2)$ is appreciable. In addition, μ is generally the most constant of the valve parameters; while g_m changes appreciably with heater voltage variations and as the valve ages. To obtain a large value of μ, V_1 and V_2 must be pentodes with the screen-to-cathode potential fixed, and arranged so that the screen currents do not flow through R_3. Fig. 2 shows a commonly used circuit. in which the anode slope resistance r_{a3} of V_3 provides a large common cathode resistance. If r_{a3} is assumed to be infinite, then:

$$\frac{M_d}{M_v} \cong \frac{\left(\mu' + 1\right)\mu'}{\mu_1' - \mu_2'}$$

where u_1' and u_2' are the inner *[control grid to screen grid]* amplification factors of the valves.

u' has been written as before for the average value of u_1' and u_2' This equation is derived on the assumption that the anode voltage does not affect either the screen or anode current of the valve and that the ratio of the anode to the screen current is constant. It predicts a slightly larger value of M_d/M_v than the exact equation. Since the inner amplification factor of a pentode is relatively low, the value of M_d/M_v is small unless u_1' and u_2' are nearly equal.

The screen to cathode potential of a pentode valve can be held constant by connecting a large capacitor between screen and cathode as shown in the circuit of Fig. 2. However, the effective resistance in the cathode circuit now becomes r_{a3} in parallel with R_4 which is comparatively small, so that M_v no longer depends mainly on the difference between u_1 and u_2 and M_i becomes relatively large.

Andrew[3] has overcome these difficulties by using the circuit of Fig. 3 for which the rejection ratio is commonly greater than 1:10,000 (–80dB). The circuit of Fig. 3 has two practical disadvantages; a floating battery is required and, unless elaborate precautions are taken, the high anode impedance of V_3 is shunted by various leakage resistances and stray capacitances. Since the heater-to-cathode impedance of most valves is relatively small, the heater supply to V_1 and V_2 must have a low resistance and, what is more difficult to achieve in practice, a low capacitance to earth.

The Feedback Circuit

Fig. 4 shows a circuit which has none of the disadvantages of that shown in Fig. 3, and in several respects gives a better performance. Its operation can be briefly explained as follows:

The essential function of a large value of cathode load [resistance] is to maintain constant the total cathode current, and hence the mean anode potential of V_1 and V_2. In the circuit of Fig. 4, this action of the common cathode resistance is

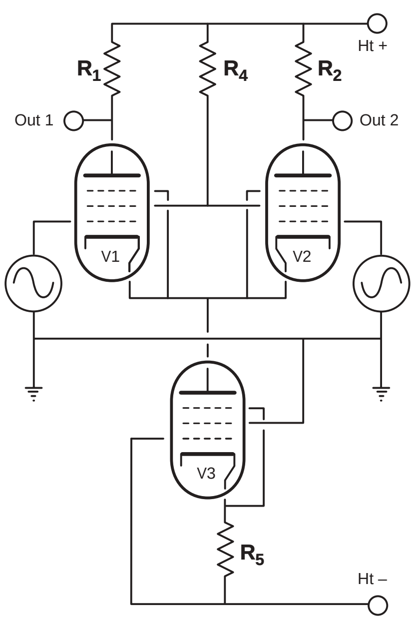

Fig. 2. A conventional differential amplifier-stage using pentodes.

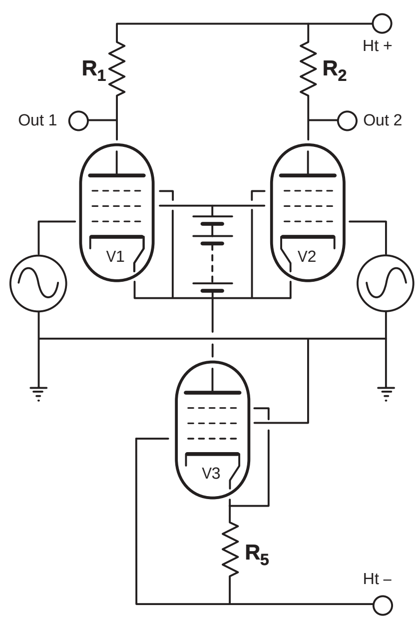

Fig. 3. Andrew's differential amplifier-stage.

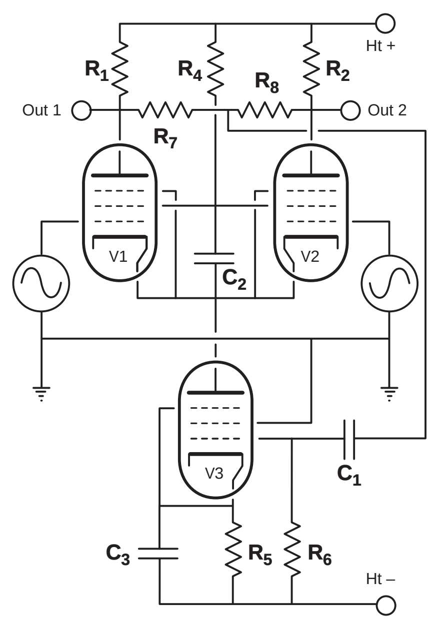

Fig. 4. A differential amplifier-stage with additional feedback.

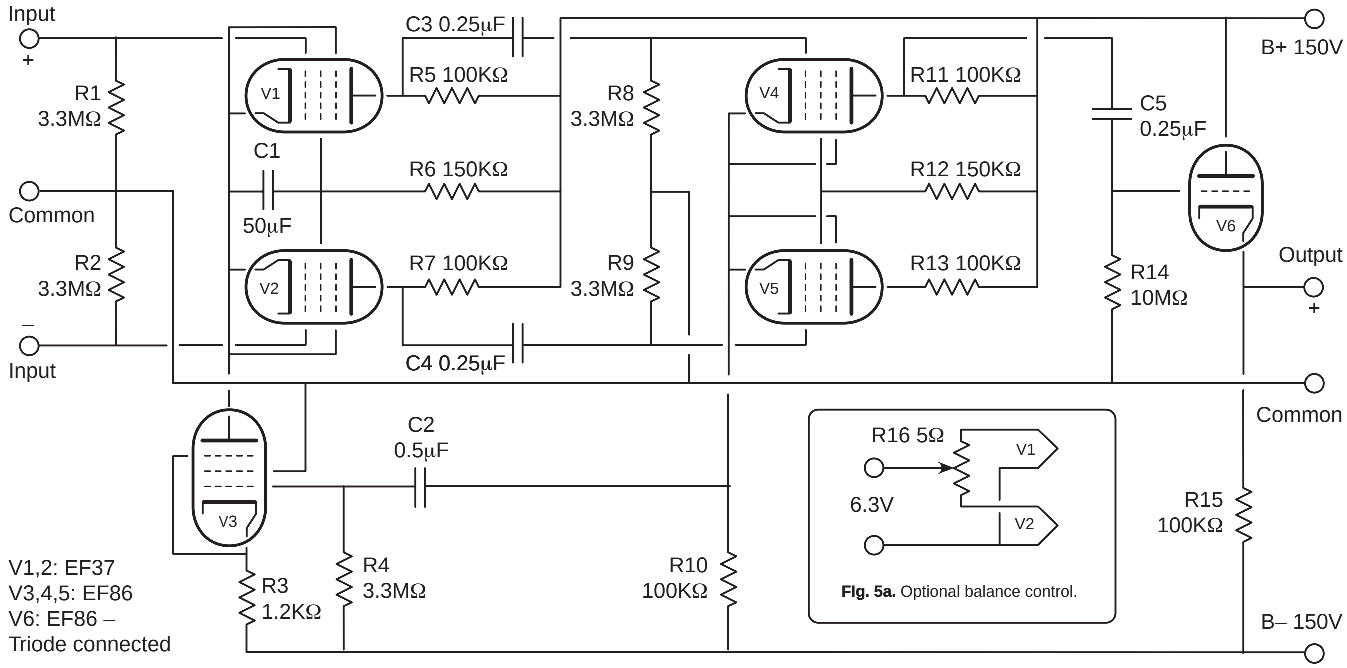

Fig. 5. A practical differential amplifier-stage.
V6: EF86 triode-connected; R_1 and R_2 are high stability carbon resistors

Table 1

Rejection Ratio	1000 to 2000	2000 to 5000	5000 to 10000	10000 to 20000	Over 20000	Total
Number of valve-pairs	1	17	2	3	5	28
Number of valve-pairs	1	11	2	3	5	21

assisted by feeding to the control grid of V_3 a potential equal to the mean anode potential of V_1 and V_2. This negative feedback loop tends to stabilize the mean anode-potential of V_1 and V_2. Although this type of feedback is not new[4], its possibilities do not appear to have been fully explored.

At frequencies where capacitive impedances can be neglected, and if $R_7 = R_8$, it can be shown that the parameters of the circuit M_d, M_i and M_v are given by Eqns.s 1, 2 and 3, except that R_3 is now replaced by $R'(1 + \frac{1}{2} g_{m3} R)$, where R' is the impedance of R_4 and r_{a3} in parallel. Since $g_{m3}R$ can readily be made 50 or 100, the effective cathode load is now very high and at least as large as r_{a3} alone. It can also be shown that if R_7 differs by only a small amount (say 20%) from R_8, the operation of the circuit is not appreciably affected. Not only is the floating battery eliminated, but any stray impedance between the cathode or screen of V_1 and V_2 and earth need be greater than R' only. In particular, an earthed heater supply can be used for V_1 and V_2.

If we make the reasonable assumptions that $(\mu + 1) \approx \mu$, $g_{m3}R \gg 2$ and $\mu_1 g_{m3} RR' \gg (R + r_a)$, then we can write:

$$\frac{M_v}{M_d} \cong \frac{g_{m1}R_1 - g_{m2}R_2}{g_m R} \bullet \frac{1}{g_m R g_{m3} R'} + \frac{\mu_1 - \mu_2}{\mu} \bullet \left(\frac{1}{\mu}\right)$$

Typically, we can take $g_m R g_{m3} R'$ as about 10,000 and μ as about 2,500 so that if $R_1 = R_2$ we find:

$$\frac{M_v}{M_d} \cong \left(\frac{1}{2500}\right)\left[\left(\Delta g_m / 4 g_m\right) + \left(\Delta\mu / \mu\right)\right]$$

where $\Delta g_m = g_{m1} - g_{m2}$ and $\Delta\mu = \mu_1 - \mu_2$. Hence it would be expected that the rejection ratio of this circuit would be at least 2,500 (68dB), and that the effect of valve ageing etc. would be small.

Practical Circuit

Fig. 5 shows a practical circuit embodying the above principles The feedback voltage for the first stage is conveniently taken from the cathodes of V_4, and V_5,. The in-phase gain of the first stage is approximately $1/60$, and of the second stage $1/3$, so that the output signal can be taken either between the anodes of V_4 and V_5 or between one anode and earth, without appreciably affecting the rejection ratio. The anti-phase gain is over 3,000 (70dB) for frequencies between 5Hz and 10kHz To check the effect of valve parameters on the discrimination ratio, a random sample of eight EF37As was taken, including both new and aged valves. The inversion gain was measured for an input of $1V_{rms}$ at 500Hz. The corresponding rejection ratios are given in line 1 of Table 1. Most of the low values of rejection-ratio arise from the presence in the sample population of one valve with anomalous characteristics. If that valve is removed from consideration, the rejection ratios become those shown in line 2. Almost half the valve-pairs then give rejection-ratios of more than 5,000 (74dB), more than a third give give ratios over 10,000 (80dB) and almost a quarter gives ratios over 20,000 (86dB).

Balance Control

Although rejection ratios adequate for most purposes can be obtained with only moderate care in selecting valves, it is sometimes desirable to have a means for increasing the rejection ratio. Only at relatively low rejection ratios (nominally <5,000) is it likely to be true that the first term on the right-hand side of Eqn. 3 is unimportant. If either R_1/R_2 or g_{m1}/g_{m2} is made variable, effectively infinite rejection-ratios can be obtained.

It is found experimentally that if the ratio R_1/R_2 is adjusted by means of a suitably connected potentiometer so as to give a very high rejection ratio, then this high ratio is not maintained when the heater voltage supplied to the amplifier is varied appreciably.

The ratio g_{m1}/g_{m2} can be varied by means of a potentiometer connected so as to vary the relative heater voltages of V_1 and V_2. By this technique, due to Aitchison[5], the characteristics of the valves can be almost exactly matched, and the balance is maintained over a wide range of supply voltages and input signals. With the control shown in Fig. 5a, the rejection ratio can be made greater than 50,000 (94dB) for most valve pairs, and this ratio does not fall below 10,000 for heater voltage changes of ±10%. At rejection ratios greater than 50,000, the output is almost wholly second harmonic.

Frequency Response

The particular amplifier shown was designed for use with a high impedance bridge network operating over the frequency range 20Hz to 10kHz and the performance figures quoted at 500Hz maintain over this range. The antiphase gain and rejection ratio fall by 3dB at about 25kHz The high frequency range could easily be extended by a factor of at least five by using valves of higher g_m (e.g Mullard EF91, EF95) and lower values of anode and screen resistors. The antiphase gain falls 3dB at about .5Hz but this figure could easily be lowered to .1Hz by simply by increasing the time-constants of the coupling networks.

The rejection ratio increases by a factor of between 2 or 3 at 10Hz but does not deteriorate by a factor of more than 10 for frequencies as low as 1Hz The exact performance depends on the balance of the valves and their associated circuits For many applications, the amplifier is required to discriminate against mains interference and a fall in rejection ratio for frequencies lower than 50Hz is unimportant. Although some improvement can be attained by increasing the various capacitors in the circuit, operation at very low frequencies is better achieved through the use of directly-coupled–amplifier techniques.

Miscellaneous Details of Performance

The rejection ratio measured at an in-phase input signal level of $1V_{rms}$ is fairly well maintained for input signals up to about $5V_{rms}$, but decreases markedly for signals greater than $10V_{rms}$. The exact figures depend to some extent on the working range over which the valves remain balanced.

The circuit is relatively insensitive to supply voltage changes, and to the impedance of and ripple on the h.t. lines, provided the impedance between the negative h.t. *[B –]* line and earth is not too large (<1KΩ). The performance given was obtained by using simple, gas-tube–stabilized power supplies.

The particular circuit described was intended for use with a high impedance bridge network. It was intended to eliminate the need for a screened and balanced transformer and to yield a single-ended output to feed a tuned amplifier and detector. For most applications there is no reason why miniature valves should not be used throughout. In the present instance EF37As were used because the top-cap grid connections were convenient in the layout adopted, and because octal-based valves were less liable to damage when repeatedly inserted and removed while the effect of valve matching was being observed.

With slight modification (e.g. increasing C_3R_{11} and C_4R_{12} to about 3sec. each) the amplifier should prove suitable for many biophysical applications.

The ability of the feedback circuit to give low in-phase gain in a single stage makes it useful in applications where a push-pull to single-ended stage is required. Such a stage, with the anti-phase gain stabilized at unity, has been designed to enable commercially made measuring instruments with a single-ended input to be used with a push-pull input. It is hoped to describe this device in a separate article.

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Inherent Feedback in Triodes

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SUMMARY—*The triode is imagined to be replaced by an infinite-impedance pentode (with its simplified anode-current expression $g_m dV_c$) with a fictitious emf in the grid circuit to represent the "back action" of the anode on the field at the cathode It is shown how this transformation makes it possible to obtain practical triode circuit formulæ from conventional feedback theory.*

IN COMPARISON WITH A PENTODE, a triode has both strong and weak points. If, in mathematical analyses, the triode be considered as an infinite-impedance pentode with negative feedback, certain of its advantages appear as direct and expected results of negative-feedback theory and in some cases a simplified analysis can result. The method is of particular interest for control circuits and output stages utilizing low-μ triodes, since the back action from the anode on the emission-controlling field at the cathode is then appreciable. Fundamentally, *this electric field action is a form of negative feedback.*‡

Fig. 1a shows a conventional pentode representative of any screen-grid, multi-electrode valve. Fig. 1b shows a conventional triode. The dynamic mutual conductance g_{md} is, for the pentode circuit:

$$g_{md} = g_m = \frac{dI_b}{dV_c} \qquad \text{.......................(1)}$$

and for the triode circuit:

$$g_{md} = \frac{r_a}{r_a + Z_b} g_m = \frac{dI_b}{dV_c} \qquad \text{.................(2)}$$

or

$$g_m = \frac{dI_b}{dV_{ce}} \qquad \text{.........................(3)}$$

where the equivalent; control voltage, *dVce*, is:

$$dV_{ce} = dV_c + \frac{dV_b}{\mu} \qquad \text{....................(4)}$$

It is seen that Eqn. 1 becomes identical with Eqn. 3 when the term dV_b/μ in Eqn. 4 tends to zero. The presence of this term may then be considered as the result of the removal of one or more shielding or screening grids in the circuit Fig. 1a. Thus, it is logical to consider dV_b/μ as a feedback voltage injected in series with dV_c and thus added to dV_c because of lack of electric shielding between the anode and the cathode. (If dV_c is positive, dV_b produces a negative term hence $dV_{ce} < dV_c$).

Eqn. 2 represents a form of the Equivalent Anode Circuit theorem. This theorem also applies to the circuit in Fig. 1c where a fictitious screen grid has been inserted between the anode and cathode to justify the transfer of the voltage dV_b from the anode circuit to the grid circuit where it appears as the fictitious voltage $dV_f = dV_b/\mu$, as required by Eqn. 4. Equivalence is now established between the circuit

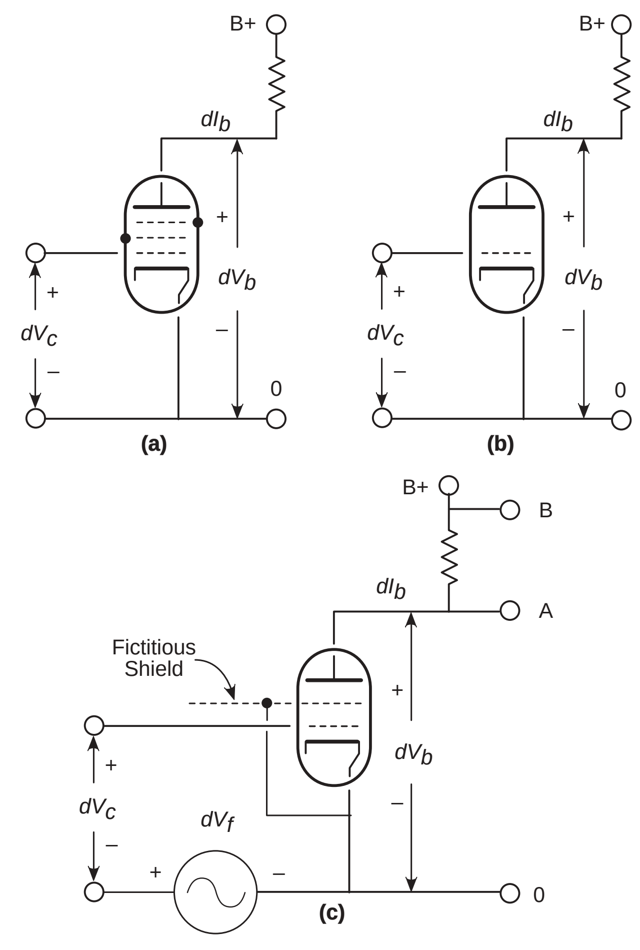

Fig. 1. Pentode and triode circuits (a) and (b) along with an equivalent (c) in which a voltage, dV_f, in the grid circuit produces the effect of a screen grid.

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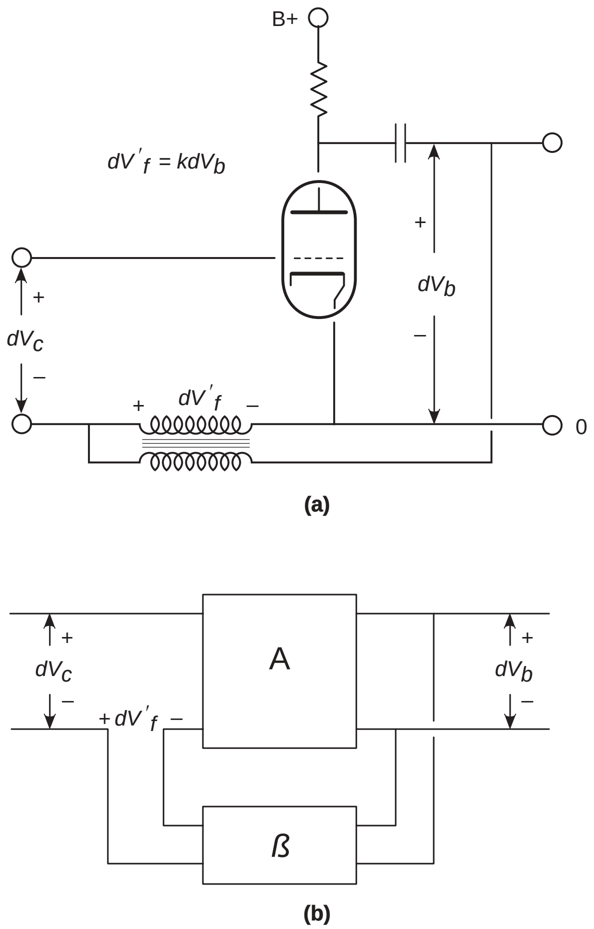

Fig. 2. (a) a circuit for obtaining the grid voltage of Fig. 1c by feedback and (b) its equivalent circuit.

in Fig. 1c and the basic circuit for voltage-controlled feedback, Fig. 2b, where A is the amplification of the triode functioning as a pentode; thus $A = -g_m Z_b$ and $ß$ is the feedback transmission coefficient $1/\mu$. The fundamental equation for a triode circuit can now be derived from conventional feedback theory. Thus, the actual amplification of the triode takes the well-known form:

$$A_a = \frac{A}{1 - \beta A} = \frac{-\mu Z_b}{r_a + Z_b} \qquad \text{(5)}$$

The combination impedance Z_{AB} seen right to left between the output terminals A, B in Fig. 1(c) can be determined if, for $dV_c = 0$, a voltage source dV_0 is applied to these output terminals sending the current dI_0 into the parallel circuit, and has the obvious form

$$Z_{AB} = \frac{dV_o}{dI_o} = \frac{r_a Z_b}{\left(r_a + Z_b\right)} \qquad \text{(6)}$$

This equation clearly expresses the reduction in output impedance due to the shunting of Z_b with the low r_a of a low-μ triode. The fictitious shield is insignificant for the above output impedance calculations; there is only one field change at the cathode, no degeneration, and $dV_f = 0$.

When external feedback is applied, it can be considered a logical addition to the already-present internal feedback, expressed by the method given above. This implies that the feedback transmission coefficient should be changed to include the externally-established coupling coefficient. As an example, a very simple external feedback circuit is shown in Fig. 2a, utilizing a transformer to inject the voltage $dV'_f = kdV_b$, where k is a constant, into the grid circuit, so that for increased negative feedback the resulting transmission coefficient becomes

$$\beta = \frac{1}{\mu} + k \qquad \text{(7)}$$

Therefore the actual amplification in this example is

$$A_a = \frac{A}{1 - \beta A} = \frac{-\mu Z_b}{r_a + \left(\mu k + 1\right) Z_b} \qquad \text{(8)}$$

Extending the example further, we may consider the transformer in the region of its upper cut-off frequency, with a peak response due to its leakage-reactance resonance. It is well known that this response curve flattens out when external negative feedback; is applied. Actually, before any external feedback is applied, *the response curve has already flattened out by the internal feedback*. If the internal feedback were removed, the response curve would be still more peaked. Thus the quality improvement due to internal feedback is of the same nature as the quality improvement due to external feedback. This line of thought pertaining to negative feedback might be useful in comparing the inferior frequency response of a pentode to the frequency response of a triode.

Considering the external application of positive feedback, it follows that we must apply a substantial amount of such feedback to a triode circuit before we have actually applied any feedback at all, in the true sense of the word. This is the feedback Eqn. 5, for if, $ßA = 0$, $A_a = A$; indicating no change due to feedback. Reversing the connections on one side of the transformer so as to provide positive feedback, it is seen that the feedback transmission is still represented by Eqn. 7, with reversed sign for k however, so that if we apply just enough positive feedback to make $k = 1/\mu$, there is no feedback at all in the circuit—$ß = 0$. The true status of feedback in a triode valve circuit is of importance when comparing different circuits with *the same amount of feedback* *applied to each circuit*. Thus, if a quantity such as reduction in noise is to be measured, first without feedback, then with specified amounts of feedback, and if the second term in Eqn. 4 is appreciable compared to the first one, it follows that equity obtains only if the “zero” feedback of the triode is compared with an amount of feedback in the pentode corresponding to the second term in Eqn. 4.

If (by changing the transformer ratio) we increase k further, the point of oscillation is reached for $ßA = 1$. Solving Eqn. 7 with $k = -k^*$ for this condition, and multiplying by A, we obtain the critical value for oscillation

$$k^* = \frac{1}{g_m Z_b} + \frac{1}{\mu} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (9)$$

This value of k is indicative of a negative resistance in the resulting loop-circuit equal to the positive loss resistance. The first term in Eqn. 7 represents the positive feedback which would be needed to make the tube oscillate if it were the equivalent of a pentode. The second term represents the additional feedback needed in a triode circuit to overcome the already-present negative feedback, which is due to back action from the anode on the electric field at the cathode.

As a mathematical criterion, $ßA = 1$ is considered a correct indication of oscillation in the above circuit. From a technical point of view, the formulæ resulting from $ßA = 1$ are not true, nor do they represent more than an approximation even when $ßA \neq 1$ or $ßA \approx 1$. The reason for this is the heavy regeneration that precedes oscillation as $ß$ is increased. The circuit is then no longer linear, and since the feedback formula is derived from Kirchoff's equations, the formula does not fully apply. While the above theory also applies to high-μ triodes, the second term of Eqn. 4 then becomes so small as to be negligible, and there is no further need to shield the anode from the cathode. Since high-μ valves are more likely to be used for high-frequency operation than low-μ valves, an entirely different shielding or screening, namely of the anode from the control grid, becomes significant. This shielding, to prevent circuit coupling (the oft-discussed Miller effect) is thoroughly treated in the literature and will not be discussed here. However, it should be noted that in cases where a low-μ triode is used at radio frequencies and if the transmission coefficient is properly modified to include the effective coupling between the anode and the grid circuits, the basic theory given above is naturally extended to include the Miller effect. Thus, one generalized feedback theory will cover both of the above-discussed gain controlling phenomena exhibited by triodes.

The original invention of the screen-grid valve in 1918 by W. Schottky, Germany, aimed at the removal of the second term in Eqn. 4 by shielding the DC field at the cathode from modulation by variations in the anode’s field.[1,9] Theoretically, *a similar improvement can be obtained by applying positive feedback to cancel the inherent negative feedback*.‡ If there existed an ideal solution to this positive feedback proposition, low-μ triode valves might today be used in many applications now employing pentodes. So far there has been no invention aiming at the elimination of the second term in Eqn. 4 that has been of any significance compared to the simple and ingenious Schottky screen-grid invention (or later beam-tube solutions). Such a development is, however, not contradictory to the basic laws of physics and may be made in the future. This is said in view of the fact that future “grid-controlled” devices, competing with the low-frequency output valve, would make use of basic principles for magnetic amplification, dielectric (ferro-electric) amplification, transistor amplification, and other amplification, where the equivalent to the second term in Eqn. 4 is either virtually non-existent or can be eliminated by methods not applicable to a vacuum tube.

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‡ Indicates my conversion to *Italic*; bp.