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The Optimised SRPP Amp (Part 1)

By Merlin Blencowe

Some deceptively-simple circuits, explained for the audio amateur.

The SRPP is a most enigmatic circuit (fig. 1), and seems to cause bewilderment and inspiration in equal measure, among many valve enthusiasts. It is frequently confused with other, vaguely similar-looking circuits, and its practical uses are not immediately obvious. This article will look briefly at the history of the circuit, its operation, and how it can be optimised for its task as an *output-transformerless power amplifier*. Oh, and also a couple of circuits that aren't SRPPs at all!

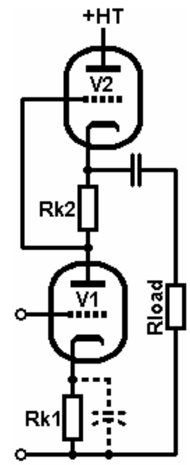

Fig. 1: The humble SRPP seems to inspire and confuse in equal measure.

What does SRPP stand for?

The first peculiarity of the SRPP is that no one seems to be sure where the name came from! The earliest reference to the circuit (that I could find) is a 1940 patent by Henry Clough of the Marconi company[1]. Most surprising is that although the patent mainly describes its use as a modulator, it does acknowledge that it could as easily be used as an audio amplifier. Yet it would be many years before it appeared in a commercial audio application.

In 1943 the SRPP was patented again in slightly different from, by Artzt at RCA[2]. Still, it does not appear to have been widely used until the advent of television, and was not given a particular name for many years, and is rarely found in radio textbooks of the time (Artzt preferred to call it a "series-balanced amplifier"[3]). Occasionally it does appear under rather generic descriptions like "two-tube series arrangement…sometimes useful as a power amplifier"[4] or "totem-pole amplifier"[5].

In 1951, Peterson and Sinclair[6] finally adapted the SRPP for audio use, ambitiously calling it a "distortionless audio amplifier" in their patent granted in 1957. In this patent they expand it into a µ-follower and finally produce what they called the *single-ended push-pull amplifier*[7], or SEPP. The SEPP, then, is related, but not identical to, the SRPP as we know it now.

By the late 1950s, textbooks on television circuits routinely referred to the SRPP as a shunt regulated amplifier[8,9], which at least explains the initials S and R. However, outside the world of television it was generally referred to as a *bootstrap amplifier*[10,11] (which also covered circuits like the µ-follower and White cathode follower) or SEPP. Of course, we now use the term SEPP for the more specific circuit arrangement by Peterson and Sinclair.

From the 1960s onwards a variety of transistor versions appeared [12,13,1415], but *still* without the name SRPP applied. However, by the 1990s we suddenly find the valve version routinely referred to as the SRPP[16,17] (although not always correctly), but exactly who first bestowed these initials remains a mystery (do *you* know?). Nevertheless, the name of *Shunt-Regulated Push-Pull* amplifier is now sure to stick, and we at least have some sense of standardisation.

Back to basics:

If audio electronics magazines are anything to go by, the operation of the SRPP seems to be routinely misunderstood as something involving a cathode follower. Therefore it might be useful to

1 Clough, N. H. (1940). Improvement in or relating to Modulator Arrangements. British patent 526418.
2 Artzt, M. (1943). Balanced Direct and Alternating Current Amplifiers. US patent 2310342.
3 Artzt, M. (1945). Survey of DC Amplifiers. *Electronics*. (August) pp212-8.
4 Valley, G. E. & Wallman, H. (1948). *Vacuum Tube Amplifiers*, pp.456-64. McGraw-Hill Book Company, Inc.
5 Millman, J. & Taub. H (1956). *Pulse and Digital Circuits*. McGraw-Hill, New York. p.100.
6 Peterson, A. P. G. and Sinclair, D. B. (1957). Distortionless Audio Amplifier. US Patent 2802907.
7 Peterson, A. P. G. & Sinclair, D. B. (1952). A Single-Ended Push-Pull Audio Amplifier. *Proceedings of the IRE* (January), pp.7-11.
8 Fink, D. G. (ed.) (1957). *Television Engineering Handbook*. McGraw-Hill, London. Ch13-11.
9 Amos, S. W and Birkinshaw, D. C. (1958). *Television Engineering, Principles and Practice*. **4**, p.241.
10 Keen, A. W. (1958). Bootstrap Circuit Technique. *Electronic and Radio Engineer*. (September). Pp. 345-354.
11 Young, J. F. (1962). Bootstrap D.C. Amplifier. *Wireless World*, November, pp.553-6.
12 Uti, S., Ueno, Y. & Kashinagi, H. (1967). Push-Pull Amplifier Operated with One Input. US Patent 3328713.
13 Kobayashi, K. (1975). Transistor Amplifying Circuit. US Patent 3890576.
14 Bellamy, P.D. & Mueller, G. (1964). Transistor Switching Circuit Responsive in Push-Pull Manner to Single-Ended Input. US Patent 3124758.
15 Butler, F. (1961). The Bootstrap Amplifier (Correspondence). *Electronic Technology*. (July) p. 267.
16 Touzelet,P. J. (1999). Theory of the SRPP Circuit. *Glass Audio*. (2) pp. 44-7.
17 Madsen, F. SRPP Power. *Sound Practices*, (17).

quickly review the operation of a simple resistance-loaded valve, so as to avoid the same confusion again.

Fig. 2*a* shows a conventional, common-cathode gain stage. The input happens to be supplied from an isolation transformer as this removes any problems with ground-referencing, and the grid is biased negative using a battery. If we input a positive pulse the valve will conduct more anode current. The voltage drop across the anode resistor Ra increases, so the signal voltage at the anode appears inverted. If we draw a Thévenin equivalent circuit for this, remembering that ground and the HT are shorted together by the large power supply smoothing capacitors, then we arrive at fig. 2*b*. We can adjust the diagram again as is *c.* to show more clearly that, as far as AC signals are concerned, Ra and r_a form a potential divider whose gain is Ra/(Ra+r_a). If we make v_{gk} equal to 1V then we can express the gain of the whole stage as:

$$A = \frac{-\mu Ra}{Ra + r_a}$$

And the output impedance is:

$$Zout = Ra \parallel r_a = \frac{Ra . r_a}{Ra + r_a}$$

Where
Ra = the anode resistor (plate resistor).
r_a = the internal anode resistance (plate resistance) of the triode.
μ = the amplification factor of the triode.

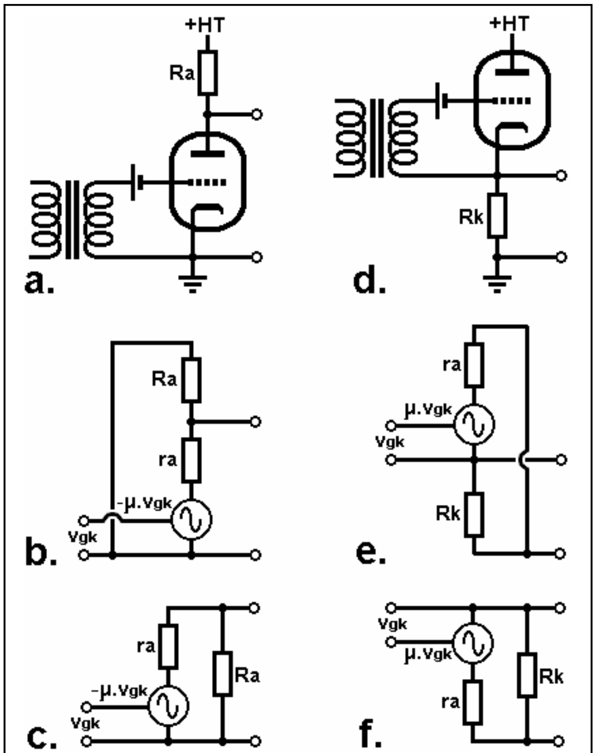

Fig. 2: **a-c:** The common-cathode gain stage and its equivalent circuit. **d-f:** The common-anode gain stage and its equivalent circuit. (Not a cathode follower in sight!)

Now consider the circuit in fig. 2d. At first glance it may look like a cathode follower, and even some experienced builders will mistake it as such. However, the secret lies in the fact that the input signal is still connected directly between *grid and cathode*, just as it was before, so all we have really moved is the load resistor, which is now called Rk. With a true cathode follower the input is applied between grid and ground and, because the cathode tries to follow the grid, the actual signal appearing between grid and cathode – which is what a valve amplifies– is very small, so we get very little voltage gain. But not so in this case! The input signal appears *directly* between grid and cathode, no matter what signal appears at the cathode. Again, if we input a positive pulse the valve will conduct more anode current, so the voltage across Rk increases and so the output voltage is not inverted this time. If we draw a Thévenin equivalent for this circuit then we arrive at fig. 2*e* and, if we adjust the diagram yet again, we come to *f.*, which is exactly the same as that in fig. 2*c* except that it is flipped! Therefore, this circuit operates *exactly* as the previous one did, except that the output signal is not inverted; it is a *common-anode gain stage*. The gain is simply:

$$A = \frac{\mu Ra}{Ra + r_a}$$ (note the minus sign is missing)

And the output impedance is the same as before.

A true SRPP is built by stacking one of these common-anode stages (V2) on top of a common-cathode stage (V1), as in fig. 1. The signal current flowing in V1 has to flow through Rk2, and so a signal voltage appears across it and forms the input signal for V2. Since Rk2 is connected between grid and cathode of V2, so the signal voltage appears directly between grid and cathode, so V2 is not a cathode follower! When we input a positive pulse to V1 it conducts more, causing the voltage across Rk2 to increase. V2 therefore receives a negative pulse and is forced to conduct less. Thus the two valves conduct in phase opposition; it is a push-pull amplifier with its own built-in phase inverter.

Why shunt regulated?

Considering the two valves are in series it might seem a little odd that it is called *shunt* regulated, which implies a parallel connection. The explanation is provided by Amos and Birkinshaw (1958): In television circuits it was often necessary to drive very capacitative loads without gross distortion or signal loss at high frequencies. However, for reasons of economy it was not desirable simply to use one or two ordinary class-A stages since they would have to idle at high current, even though the actual current demands might be only small most of the time. The solution to this is shown in fig. 3*a* (for AC only), and consists of an ordinary, low current amplifier stage V1 which handles the signal most of the time, and a sort of auxiliary valve V2 which provides the heavy current demands, when necessary. By placing a current-sensing resistor R in series with the load Z, if the load current increases (i.e., the signal frequency is increased) the voltage drop across R also increases, and if this signal is delivered to V2 it will in turn deliver

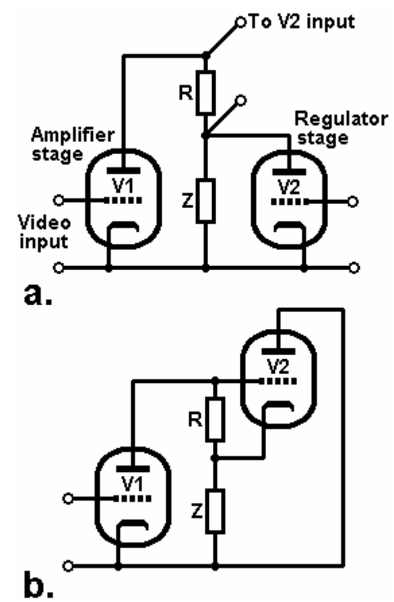

Fig. 3: Shunt regulation of a video signal (ignoring DC conditions). After Amos and Birkinshaw, 1958, p240.

the extra load-current demand. It is obvious that V2 is in parallel with the load and maintains or regulates the signal against a falling load impedance at increasing frequency. From this principle it is a short step to the SRPP in fig. 3*b*.

An important point to recognise about the SRPP is that it can only operate in Class A, because the valves are in series. If one valve cuts off then the remaining valve can no longer conduct either. The White cathode follower is also a type of strictly-class-A shunt-regulated amplifier, and was similarly popular in television and computing circuits.

When is the SRPP not an SRPP?

Whether or not the SRPP really *is* an SRPP depends on the load. If there is no load on the valve (e.g., it is DC coupled to another stage) then there is nowhere for current to go but straight through both valves, and so the stage is entirely single-ended, as in fig. 4*a*. This is *not* an SRPP but a common-cathode gain stage with an active load, and the maximum peak current flow is then equal to the quiescent current, as it always is with single ended operation. Such a circuit is obviously intended only for voltage gain, and is not widely used*.

Nonetheless, if we move the output to the anode of V1 then we obtain some useful properties∫. V2 appears to multiply or bootstrap the value of Rk2 by a factor of μ+1, so the total load on the lower valve becomes $r_a + Rk2(\mu+1)$ (which is much too low to be considered a constant-current source).

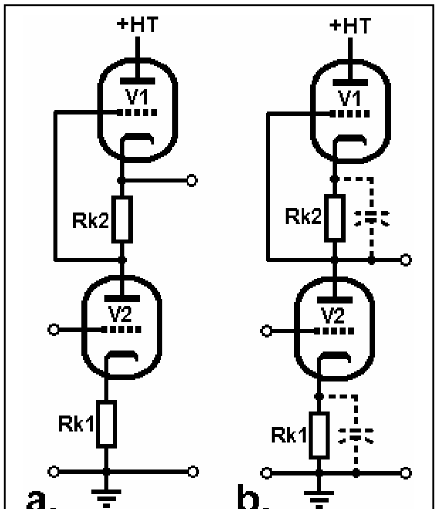

Fig. 4: **a.** With an infinite load the circuit is not an SRPP at all, but an actively loaded triode. **b.** By moving the output to the anode of V1 we obtain the *half-μ stage*. Bypassing both resistors reduces the output impedance, but distortion will be greater.

If both resistors are bypassed then the voltage gain would be:

$$A_{(bypassed)} = -\mu_1 . \frac{ra_2}{ra_1 + ra_2}$$

Or if they are unbypassed then it is:

$$A_{(unbypassed)} = -\mu_1 . \frac{ra_2 + Rk2(\mu_2+1)}{\left[ra_1 + Rk1(\mu_1+1)\right] + \left[ra_2 + Rk2(\mu_2+1)\right]}$$

Where:
Subscipt 1 indicates the lower triode.
Subscript 2 indicates the upper triode.

However, the magic really happens when we use identical triodes and also make Rk1 equal to Rk2, because the above equations then simplify to:

$$A = -\frac{\mu}{2}$$

So bypassing both resistors does not affect the gain; only the output impedance is reduced (though the distortion will be much greater). For the bypassed version:

$$Zout_{(bypassed)} = \frac{ra}{2}$$

And for the unbypassed version:

$$Zout_{(unbypassed)} = \frac{ra + Rk(\mu+1)}{2}$$

Consequently, this circuit – which we might call a 'half-μ' stage– is very useful indeed, since the gain and Miller capacitance remain quite constant despite aging of the valves, or the use of valves of varying provenance, because μ is the most consistent and stable parameter of a triode. It should also be quite immune to changes in heater power and supply voltage. On the other hand, the main disadvantage is that the PSRR is only 6dB.

The optimised SRPP:

If we attach a finite load to fig. 4*a* then we provide an extra path for current to flow. When the lower triode is driven to cut-off, V2 reaches its maximum conduction and this current flows into the load. When V1 then conducts fully, V2 reaches its minimum conduction and current flows out of the load and down the lower triode. We see that current is actively pushed into, and pulled out of the load; the load receives the difference in current draw between the two valves, hence it is push-pull operation and we have a true SRPP. Obviously, if the load impedance is very high –like a 1MΩ resistor– then the

* Interestingly, with a critical value for Rk2 it is possible to make V1 act as a genuine constant-current source, but that is another story.

∫ It is also interesting that a patent for this circuit was applied for just fifty days before Artzt's applied for his SRPP patent. See:
Bowman, J. R. (1940). Amplifier. US patent 2326614.

current delivered into it is tiny and we have a pretty pathetic example of push-pull operation, but it *is* push-pull nonetheless.

In old TV circuits it was usual for V2 to deliver more current to the load than V1. However, for audio purposes we are more interested in making both valves contribute *equally* to the load current, and to maximise that current, and this is what we will call the 'optimised SRPP'. It has already been observed by a few authors[18,19] that a deep null in the THD can be obtained with the right combination of Rk and R_{load}, and the values correspond (almost) exactly with those of the optimised circuit. This is unsurprising, for a perfectly balanced push-pull amplifier will cancel all even-order harmonics, resulting in minimum THD. However, the SRPP does not have truly push-pull input signals, since the input signal to the upper triode has already been amplified by the lower, and so will contain some extra distortion. As a result, the value of load which gives the deepest distortion cancellation is usually 10-20% lower than the value found by calculation, and will also vary rather unpredictably according to the manufacturing variations between samples. However, it is worth nothing that it is better to use a load which is slightly too large than one which is too small, as the distortion rapidly increases a below the critical value.

Next time…

If you were hoping to see a little more of the SRPP in this article (and not the half-μ amplifier) then I must confess to a little showmanship; always leave them wanting more. However, it in part 2 of this I promise to satisfy your expectations and show how to calculate the optimum resistor values for the SRPP, and look at a couple of the common circuit variations and their uses.

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[18] Evers, M. V. (1996). Distortion Minima Loading of the SRPP. *Sound Practices*. (13) pp. 40-1.
[19] Perugini, S. (2000). SRPP's Harmonic Cancellation Capabilities. *Glass Audio*. (3) pp. 42-51.

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The Optimised SRPP Amp (Part 2)

By Merlin Blencowe

With some theory behind us, it is time to really take the SRPP apart.

In the previous part of this article I looked at the history of the SRPP (as quickly as I could), and covered the half-μ stage, which is often confused with the SRPP. I also hinted that, as a small power amplifier, there must be some optimum load impedance for the SRPP, one which gives (almost) the lowest distortion. It is now time to find what that load is, and along the way I will do my best to use as little mathematics as possible!

The optimised SRPP again:

For audio purposes we are most interested in making both valves in the SRPP contribute *equally* to the load current, and also to maximise that current, and this is what we will call the 'optimised SRPP'. It is at or close to this condition that we obtain the lowest distortion. Genuinely obtaining the absolute minimum distortion requires a distortion analyser, as the critical values of resistance vary too much with manufacturing tolerances to be attacked on paper alone. However, even if you don't have such test equipment, the values calculated using the formulae given here will always bring you *close* to the ideal, and will err on the safe side of less distortion rather than more. All you need is the HT voltage (i.e., the power supply voltage or B+), and also the published figures for the internal anode resistance (plate resistance) r_a, and the amplification factor μ.

The bypassed, optimised SRPP:

Let us first examine the textbook version of the SRPP (fig. 5*a*) when the lower bias resistor Rk1 is bypassed by a large capacitor. From basic theory we know the quiescent current through a triode is:

$$I_o = \frac{HT}{r_a} + g_m.v_{gk}$$

Where:

g_m = the transconductance of the triode.

v_{gk} = the grid-to-cathode or bias voltage.

For a cathode-biased stage, v_{gk} is determined by $-I_oRk$, giving:

$$I_o = \frac{HT}{r_a} + g_m(-I_oRk)$$

If we now solve for I_o we obtain:

$$I_o = \frac{HT}{r_a(g_m.Rk + 1)}$$

And since $r_ag_m = \mu$, this may be written as:

$$I_o = \frac{HT}{r_a + \mu Rk}$$

The simple SRPP (without the additional resistor Ra in fig. 5) is two such triode amplifiers connected in series, so the denominator of the above is doubled to:

$$I_o = \frac{HT}{2r_a + 2\mu Rk} \quad \text{.....................................I}$$

The circuit is balanced when the quiescent current is half the maximum or peak current which could ever flow through the triodes. If we split the circuit into its two halves as in fig. 5*b* then we can see more easily what the maximum peak currents are. If the HT is shared equally between the two triodes then for the lower triode, assuming Rk1 is bypassed, the maximum current is $½HT/(r_a+R_{load}+Rk2)$ while for the upper triode it is $½HT/(r_a+R_{load})$. Since they are not quite balanced we should add the additional resistor Ra, equal to Rk2, in series with the upper triode as shown, in order to equalize these currents. In other words, adding Ra equalizes the effective transconductance of the two devices (this was also included in Artzt's patent).

If we say that Rk1 = Rk2 = Ra = **Rk**, the peak current for either valve becomes:

$$I_{pk} = \frac{HT}{2(r_a + R_{load} + Rk)}$$

So for proper balance the quiescent current must be half the peak current value:

$$I_o = \frac{HT}{4(r_a + R_{load} + Rk)} \quad \text{..II}$$

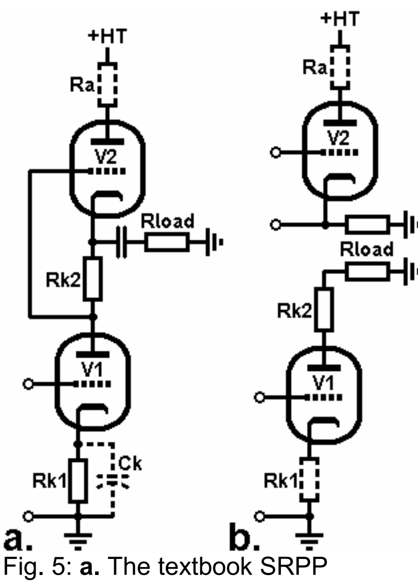

Fig. 5: **a.** The textbook SRPP amplifier. **b.** Breaking it in two to see more clearly the resistances in the signal path.

How can we force the quiescent current to equal this ideal value, for a given load impedance? There must be an optimum value of bias resistance Rk which will achieve this. This value is easily found by equating I and II (and remembering to add the extra resistor Ra). Thus we find the optimum value for the bias resistors:

$$4(r_a + R_{load} + Rk) = 2r_a + 2\mu Rk + Rk$$

$$2r_a + 2R_{load} = r_a + \mu Rk - 1.5Rk$$

$$r_a + 2R_{load} = (\mu - 1.5)Rk$$

$$Rk_{optimum} = \frac{2R_{load} + r_a}{\mu - 1.5}$$

Or if we prefer to know the optimum load for a given Rk:

$$R_{load} = \frac{Rk(\mu - 1.5) - r_a}{2}$$

However, with high-μ valves Rk will probably be very small compared to r_a, so the imbalance which occurs due to leaving out Ra is quite small. If we do chose to omit Ra then the above can be simplified to:

$$Rk_{optimum} \approx \frac{2R_{load} + r_a}{\mu} = \frac{2R_{load}}{\mu} + \frac{1}{g_m}$$

with only a modest increase in distortion*.

Applying Thévenin's theorem (which I won't prove here), we find the voltage gain of the bypassed SRPP:

$$A_{(Rk1\ bypassed)} = -\mu_1 . \frac{R_{load}(ra_2 + Ra + \mu_2 Rk)}{(ra_1 + Rk)(ra_2 + Ra) + R_{load}\left[ra_1 + ra_2 + Ra + Rk(\mu_2 + 1)\right]}$$

But for the optimised circuit where both triodes are identical and Rk2 = Ra = Rk, this simplifies to:

$$A_{(Rk1bypassed)} = -\mu . \frac{R_{load}\left[r_a + Rk(\mu + 1)\right]}{(r_a + Rk)^2 + R_{load}\left[2r_a + Rk(\mu + 1)\right]}$$

The output impedance is:

$$Zout_{(Rk1bypassed)} = \frac{(ra_2 + Ra)(ra_1 + Rk)}{ra_1 + ra_2 + Ra + Rk(\mu_2 + 1)}$$

Which again simplifies to:

$$Zout_{(Rk1bypassed)} = \frac{(r_a + Rk)^2}{2r_a + Rk(\mu + 2)}$$

From the foregoing we see that there is always an optimum value for Rk for any value of R_{load}. If we change the load we must change Rk too, or linearity will probably suffer (too many authors try to ignore the load when analyzing the circuit, which leads to quite misleading results).
It is therefore a mistake to use an SRPP as the output stage or line-driver of a preamp, for example, because we can't be sure what it will be plugged in to and hence what R_{load} might be. It could be as high as 1MΩ if it gets partnered with a valve power amp, or as low as 10kΩ with a transistor amp, and we can hardly optimise for every eventuality. The SRPP can only be designed properly if we know what the load impedance actually is, and treat the whole thing as a unified system (remember, an SRPP with no load is not an SRPP at all!).

However, if μ >> 1, and $R_{load} << r_a$, then $Rk_{optimum}$ approaches r_a/μ or $1/g_m$, and variations in load impedance will have much less impact. Unusually, then, the SRPP is best suited to *low* impedance loads. It is fundamentally a 'small power amplifier', and ideal applications include driving dynamic headphones, which vary from around 15Ω to 300Ω and require only a few milliwatts of power. One early example of a commercial audio application was in the Philips AG2126 "Magic Box" record player[1], where a pair of triode-connected UL84's were arranged as a bypassed SRPP to drive high impedance loudspeakers.

If we have optimised our SRPP properly then we now know that the peak load current is equal to twice the quiescent current, and from $P = I^2R$ the maximum output power must be:

$$P_{max} = R_{load}\left(\frac{HT}{\sqrt{2}\left(r_a + \mu Rk + \frac{Ra}{2}\right)}\right)^2$$

What may not be immediately obvious is that *absolute maximum power* from the SRPP obtains when the load impedance is equal to r_a so, in theory, it should be possible not only to optimise the circuit resistances but also to choose the optimum triode for the job! All we have to do is find one whose r_a is equal to the impedance of thing we wish to drive, and which can also handle the dissipation of course. For example, if the load is 10kΩ then an ECC82 / 12AU7 comes pretty close with $r_a \approx 10k\Omega$.

Less distortion *and* more power?

Hopefully the preceding section illustrates how the optimisation process works. However, we are more likely to find SRPPs where all resistors are unbypassed as in fig. 6. Not only does this avoid the use of another large, electrolytic capacitor, which are universally hated by audiophiles, but it also reduces distortion and even facilitates *more* output power! The last point may seem counter-intuitive, but will become clearer later. The disadvantages of this arrangement are that the voltage gain and PSRR will be lower, and the output impedance will be much higher. On the other hand, since we are designing the circuit specifically to deliver power into a known,

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* This is the same formula as presented in: Millman, J. & Taub. H (1956). *Pulse and Digital Circuits*. McGraw-Hill, New York. p100.

[1] (1956) Radio Show Review. *Wireless World*. (October) pp. 475-6.

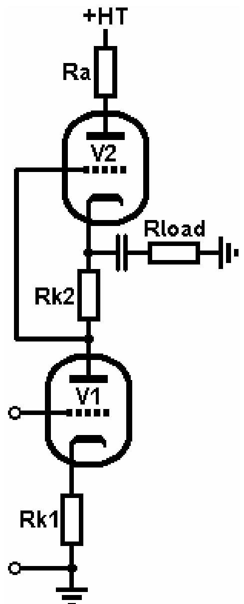

Fig. 6: With this unbypassed version of the SRPP, Rk1 = Rk2 = Rk, and the optimum value is:

$$Rk_{optimum} = \frac{2R_{load} + r_a}{\mu - 1}$$

Ra should be twice this value.

fixed load impedance, the output impedance should be of only marginal concern.

Another reason why we might not wish to bypass Rk1 is that, when it *is* bypassed, the even-harmonic distortion of both triodes will largely cancel out leaving only low-level odd harmonics, which is what we would expect from any balanced push-pull amplifier. However, by leaving R1 unbypassed, current feedback will greatly reduce the distortion generated in the lower triode as compared with the upper triode. The total distortion of the circuit will then be dominated by the second-harmonic distortion of the upper triode, so retaining that 'warmth of tone' that we expect from triodes.

The unbypassed, optimised SRPP:

If Rk1 is unbypassed then Ra should be made equal to twice Rk2 for proper balance although, again, the error is small with high-μ triodes. If the resistor is included, though, then the optimum value for Rk becomes:

$$Rk_{optimum} = \frac{2R_{load} + r_a}{\mu - 1}$$

And the quiescent current will be:

$$I_o = \frac{HT}{2\left[r_a + Rk(\mu + 2)\right]}$$

The voltage gain is:

$$A_{(Rk1\ unbypassed)} = -\mu_1 . \frac{R_{load}(ra_2 + Ra + \mu_2 Rk2)}{\left[ra_1 + Rk2 + Rk1(\mu_1 + 1)\right](ra_2 + Ra) + R_{load}\left[ra_1 + ra_2 + Ra + Rk1(\mu_1 + 1) + Rk2(\mu_2 + 1)\right]}$$

But if you don't like the look of that then for the optimised circuit it simplifies to:

$$A_{(Rk1\ unbypassed)} = -\mu . \frac{R_{load}\left[r_a + Rk(\mu + 2)\right]}{\left[r_a + Rk(\mu + 2)\right](r_a + 2Rk) + R_{load}\left[2r_a + 2Rk(\mu + 2)\right]}$$

The output impedance will be much higher than for fig. 5, being:

$$Zout_{(Rk1 unbypassed)} = \frac{(r_a + 2Rk)\left[r_a + Rk(\mu + 2)\right]}{2r_a + 2Rk(\mu + 2)}$$

And the maximum output power is:

$$P_{max} = R_{load}\left(\frac{HT}{\sqrt{2}\left[r_a + Rk(\mu + 1)\right]}\right)^2$$

The above equation still does not give anything away, but when we observe that the optimum value for Rk is *less* than for the bypassed version, we will find that the available output power is actually greater! However, before we deride the bypassed version for being a weakling it must be pointed out that the difference in power is quite negligible for triodes with μ greater than about 10.

The SRPP+

In some cases we will find that the optimum value of Rk is too small to bias the valves properly without causing them to over heat. If we can't increase the load impedance and we don't want to use a more powerful pair of valves, we can instead bias the valves to any value of current we like, and restore balance by connecting the load to a tapping point on Rk2 instead, or in other words, split Rk2 into two parts, R1and R2. This gives us the circuit in fig. 7, which was rediscovered recently by Broskie[2] and to which he has given the name SRPP+.

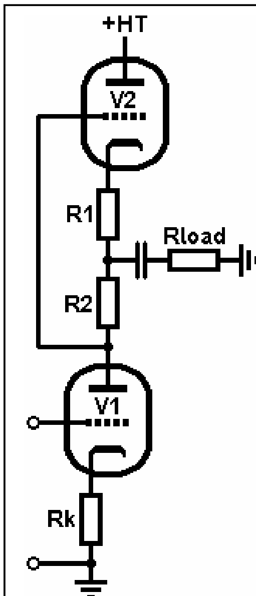

Fig. 7: The SRPP+ circuit can be used when the optimum value for Rk is too small to bias the valves properly.

The supply voltage and quiescent current can be chosen first, according to the desired dissipation in the valves (which is approximately $P = HT / 2 \times I_o$). Rather than bore the reader with the same mathematical treatment again, we will quickly consider only the unbypassed version of the circuit using high-μ triodes, as this allows some simplifications to be made.

Having decided what quiescent current I_o is allowable, the necessary value for Rk to bias the triodes to this current is:

$$Rk = \frac{HT/2I_o - r_a}{\mu}$$

[2] http://www.tubecad.com/2009/09/blog0172.htm

And R1 + R2 must also equal this value. The purpose of R1 is to deliberately degenerate the transconductance of the upper triode so that it becomes equal to 1/R2, at which point the maximum peak current in the upper triode will (more or less) match the lower triode, and will be equal to twice the quiescent current, just like the other optimised SRPPs.

The value of R1 is found to be:

$$R1 = \frac{{HT}/{4I_o} - r_a - R_{load}}{\mu}$$

And R2 is simply Rk – R1, or if we prefer to find R2 directly:

$$R2 = \frac{{HT}/{4I_o} + R_{load}}{\mu}$$

which is slightly simpler than the previous equation. If we find ourselves with a negative value for either of these resistors then it indicates that the load impedance is simply too large for the circuit to work. This demonstrates yet again that the SRPP actually prefers low-impedance loads!
Strictly, for AC balance, an extra current-balancing resistor Ra should be added, having a value of Rk + R2 – R1. However, the error caused by omitting this extra resistor is even smaller than for the ordinary SRPP. This is fortunate, since leaving it out considerably simplifies the maths.

Unfortunately, the exact expressions for gain and output impedance of this circuit and painfully long, and liable to make half the readership faint at the very sight of the SRPP+. Therefore I shall be diplomatic and say that the gain,output impedance and output power are *very nearly* the same as for the ordinary, unbypassed SRPP, with the same value Rk and R_{load}.

Conclusions:

Although I have not been exhaustive, hopefully this article will have gone some way to de-mystifying the operation of the SRPP for some readers, and illustrated what the SRPP does best. It is a genuine push-pull amplifier, and is best suited to small *power* applications, and will drive surprisingly heavy loads. Dynamic headphones (normally around 15Ω to 300Ω) are one ideal application, and a pair of ECC82 / 12AU7, ECC88 / 6DJ8 or 6SN7 are more than capable of making a good SRPP headphone amplifier which will drive most types of headphones, even cheap, low impedance earbuds!

The textbook version of the SRPP, where Rk1 is bypassed, offers about twice the gain and half the output impedance compared to the equivalent unbypassed version, but the available output power is actually slightly less. So unless we really need the extra gain, most designers will choose the unbypassed version. Not only does it offer less distortion, but the higher output impedance will allow the use of a smaller output coupling capacitor. For example, an ECC88 / 6DJ8 driving a 15Ω headphone would have an output impedance of about 910Ω bypassed, or 1380Ω unbypassed. For a cut-off frequency of 20Hz the bypassed version would need a 9μF coupling capacitor, whereas the unbypassed version would need only 6μF, making it much easier to use a good-quality, non-electrolytic capacitor. Yet from the same 200V power supply both would deliver a maximum of 10mW, with surprisingly good linearity! This illustrates rather nicely how the output impedance of the SRPP is largely incidental, because the whole circuit is designed with a specific load impedance in mind. (By the way, if you actually feel like building such a circuit, the unbypassed version would need Rk ≈ 82Ω, giving a quiescent current of 19mA. The gain would be about 0.2, so an input voltage of about 2Vrms would be needed for full output.)
What the SRPP is *not* is an all-purpose line driver, because in such cases the load impedance could vary over a very wide range. A load impedance which is somewhat higher than the design value might not be too bad, but one which is lower would be disastrous. A cathode follower would be a much more sensible choice for this position.
On the other hand, if all we need is *voltage* gain, and the load impedance is fairly high (a few hundred kilo-ohms or more, say), then the half-μ stage is a much more sensible choice than the SRPP, and offers extremely consistent results. It could make an ideal input stage for an MM phono stage, for example, where predictable gain and input capacitance are essential.
And if this article has done nothing else, then I hope that is will at least make you look a little closer before giving just any totem-pole stage the grand title of *SRPP*!