The Jean Hiraga Nemesis Quadrige

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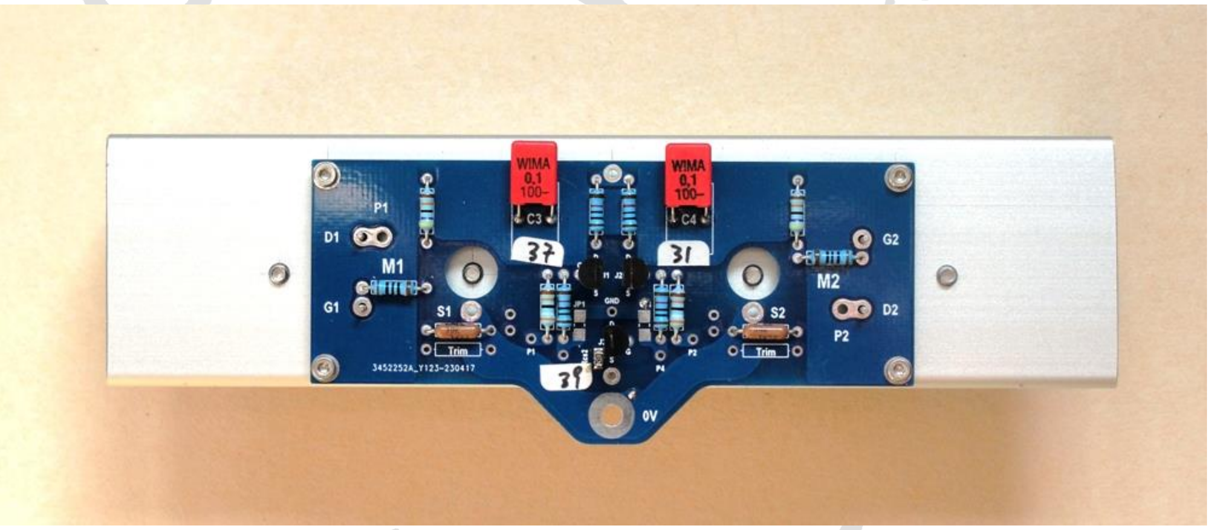


Why the Buzz ?

I am never a tube guy, but someone was buzzing about the BUZ901 [1] recently, which has been very popular in hybrid tube circuits. John Broskie, for example, has published a whole bunch of circuits based on those [2]. This reminds me also of Susan Parker's Zeus, which caused quite a stir almost 20 years ago [3].

The Zeus and the BUZ

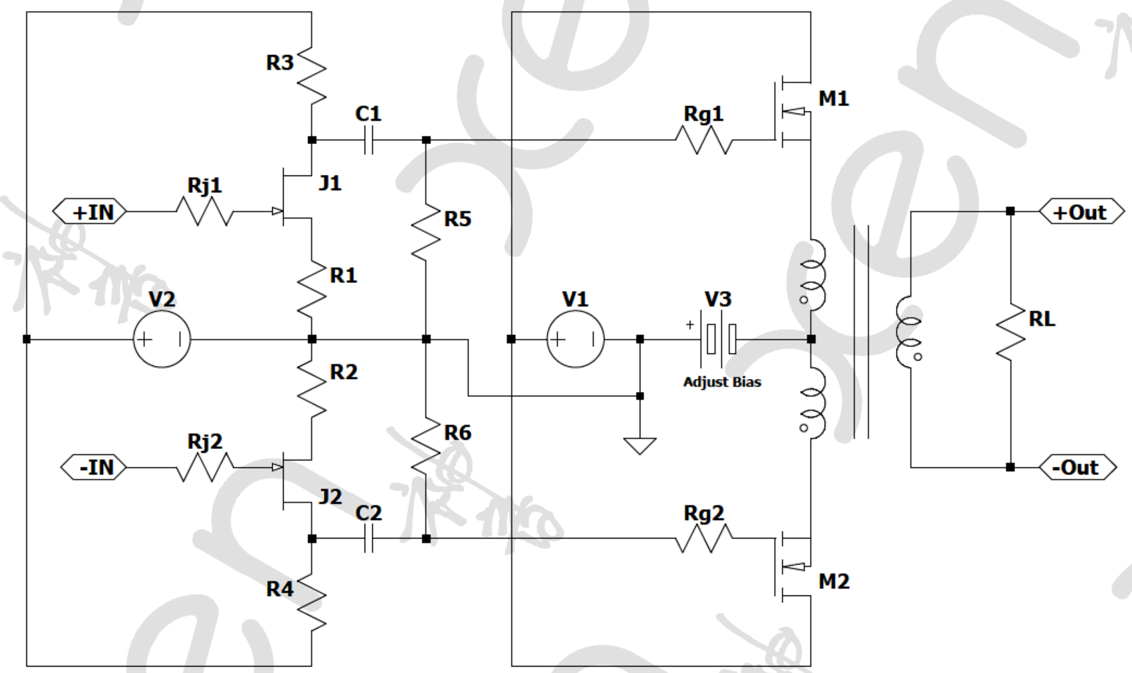
It was a wake-up call that I still have 25 pieces of BUZ900 lying in the drawer for years. Curve traced, and they can make up 5 quads or twin-pairs. So better find some nice application to put them to good use. Rather than some SE circuit, such as a tube-like SRPP, or those with a choke load at the drain [4], I like Susan's super-simple Zeus circuit.

Actually, the BUZ900 fits this quite well. It has negative tempco above 100mA, and is therefore self-stabilising. It has near constant capacitances with varying Vds, and is flat over the audio band. What it does not have is large transconductance. But this is compensated for by using the step-down output transformer (of e.g. 4:1 into 8 ohm). It means more voltage swing, but that is not a problem with its 160V Vds rating, and low capacitances. And it is in follower mode in Susan's Zeus.

The Sowter output transformer specified by Susan is rather expensive. But she has shown success using standard power-supply toroidal transformers from RS [5].

If the maximum output power is limited to say 24W into 8ohm, or +/-19.5V, this means +/-78V at the balanced input to the BUZ gates, well within its 160V Vds rating. But how to generate this +/-78V balanced input from line level ? Both Broskie and Parker were using input transformers. But with a required gain of 14x at line level inputs, the transformer primary needs to be driven by a low-impedance source to provide sufficient bandwidth. One solution can be a simple FET common-source gain stage instead. This, though, needs coupling caps and has poor PSRR, the high input impedance and high CMRR of the Zeus make life easier than otherwise.

For this solution, one needs to use a high voltage FET. The 2SK373 is the only JFET that can sustain 100V and with an Idss of about 4mA. But for high bandwidth and low distortion, a depletion power MOSFET such as IXPT01N100D can cope with much higher bias and hence lower output impedance and higher bandwidth.

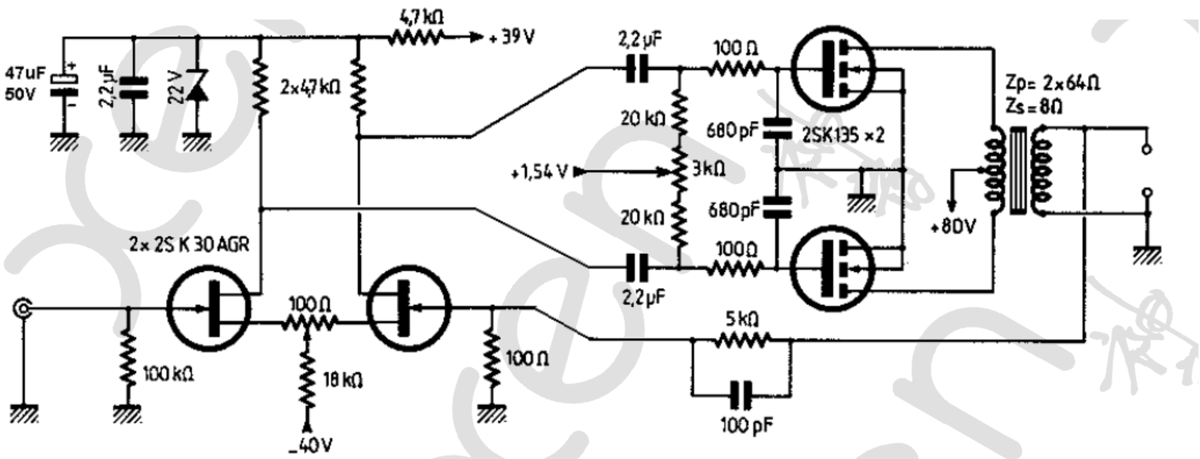


Quite a good solution actually, except for the expense of a high-voltage supply required for the MOSFET frontend. Can we not use the BUZ in common source instead to get some additional gain from the power stage ?

The Nemesis Quadrige

Jean Hiraga's single-ended Nemesis, based on the 2SK135, is widely published with a number of rebuilds, also by Nelson Pass. The BUZ900 is supposedly an improved replacement of the 2SK135. But the single-ended Nemesis' high bias current automatically means large and expensive EI transformers. A push-pull balanced circuit, on the other hand, means the magnetic fields of the opposing DC current cancels in the transformer core, allowing smaller and a larger choice of transformers, as in Susan's Zeus.

But there is actually a push-pull version of the Nemesis, known as the Nemesis Quadrige [6,7], using a pair of lateral FETs in common source with the centre tapped transformer primaries at the drain. This has an additional differential-pair front end using 2 matched 2SK30AGR's, working at normal preamp voltages. So this is the one to try out first. The transformer is essentially the same as for the Zeus with 4:1 ratio, so one can use the same to try out both, if so wish. Only the front end needs to be different.



Hiraga quoted the output power to be 15W Class A, and a steadily rising distortion with the increase in power output, between 0.01% (0.1W) and about 2% (60W Class AB), with only 6dB of feedback. Bandwidth was between 20 Hz and 75 kHz at -3 dB.

Adjustments to the Original Circuit

A Spice simulation was quickly established to enable a better understanding of the original circuit. This proved to be already very close to optimum. Nevertheless, there are a few minor areas where changes were made :

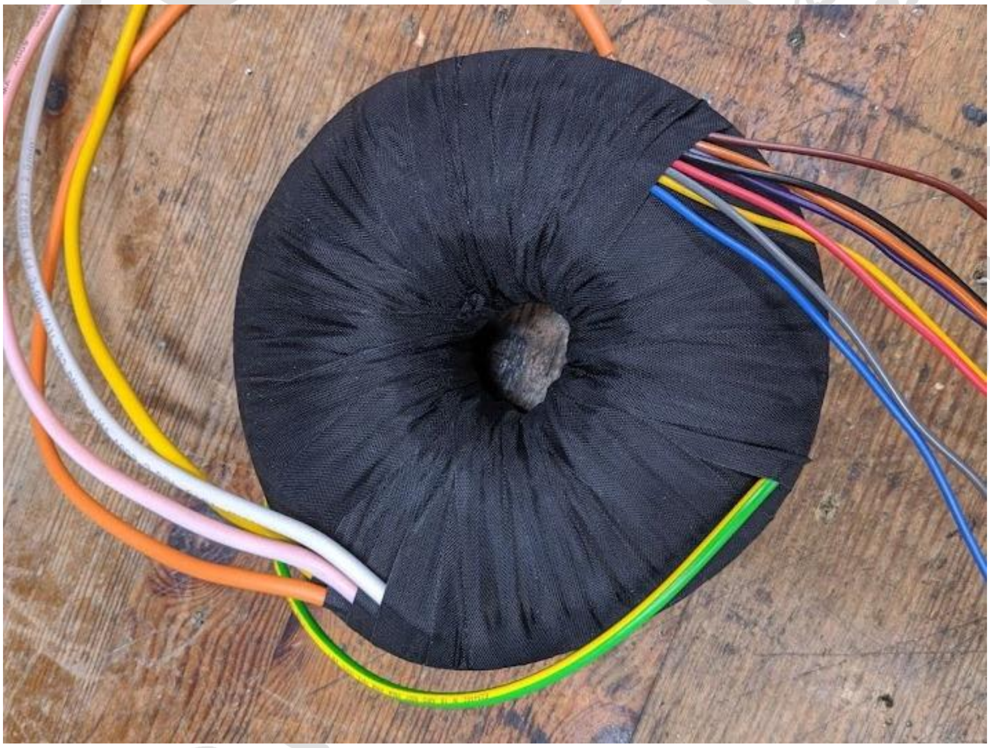
1. Instead of a separate (~1.5V) voltage source to bias the MOSFET gates, two separate potential dividers are used instead, powered from the positive supply of the frontend. This not only saves an extra power supply, but also allow individual bias voltage adjustment to allow for any slight differences in MOSFET Vgs.
2. The positive and negative rails of the frontend circuit is changed to +/-17V, which allows a wider selection of low-noise regulators.
3. The input differential pair is biased by a 2.3mA JFET current source, instead of a fixed resistor.
4. The drain resistors are increased to 10k, and the feedback network to 100k:10k, so as to set the NFB to about 30dB, and a closed loop gain of 10. It also allows the circuit to operate as fully balanced. But it is easy enough to restore to Hiraga's original values. The feedback capacitor is reduced to 10pF.
5. Additional source resistors at the MOSFETs improve bias stability further and enable a simple bias-balancing circuit to minimise net DC current in the transformer core.
6. The MOSFET gate stoppers are changed to 1k, and the gate-source capacitors changed to gate-drain and to 68p, as they can then be directly soldered to the MOSFET pins These are required for stability in the prototype test.

By the way, as transformers cannot couple DC voltages, the speaker will not get any DC, and is inherently safe. No speaker protection circuit is required.

Power Supplies

Following the advice of Jan Didden [8], a Connex SMPS300RS 90V was chosen for the MOSFET power supply. This has a regulated output voltage, as well as more than sufficient current to power 2 channels. This supply is further cleaned of any HF noise with a capacitor multiplier, using a single IRFP240 as pass device. The frontend supply can be ordinary linear supply using transformers, but one can also make use of the +/-19V auxiliary supply of the SMPS with additional LDO regulators.

The Output Transformer



Another critical component is the output transformer. The required ratio is 64R+64R:8R, or 2+2:1 in turns ratio. This is a lot lower in ratio than commercial push-pull output transformers meant for tubes. In fact, only one "standard" product called PT-87 from a Chinese source comes close. This is an EI transformer with 4 coils of 2:2:1:1, and can be wired as 32R+32R:4R. With the paralleled secondary rated at 2A total, this can be used in the circuit, but will not make full use of the potential output power.

Another candidate could be the Lundahl LL2410. This has a total of 12 windings, namely 2+2+2+2 and 1+1+1+1+1+1+1+1. So if we parallel 2 groups of [2+2 : 1//1+1//1], we should get the desired turn-ratio. The internal resistances are very low, but there is not much information, other than that it is a "general purpose 100V loudspeaker transformer well suited for applications with power levels from 250W and up, line voltage from 70V to 140V". Allowable coil voltages are sufficient for the purpose.

The Lundahl is quite expensive at 200€ apiece. We wanted to follow the approach taken by Susan Parker in experimenting with mains toroidal transformers. Examples of mains toroids used as ESL step-up transformers can be found on the internet. And there are toroidal transformers from Menno v.d. Veen and Toroidy for push-pull tube amps. It must be possible to use toroidal transformers in this application.

A famous European transformer designer gave us a hint to specify the transformer as a mains transformer. Typically, this has 2x 115V primaries, and as such the secondary would have to be 58V (for 8 ohm) to give the correct turn ratio. But to get down to 0.2Ω resistance, assuming 5% voltage regulation, the transformer would have to be rated at 800VA, which is overly large.

What if we specify 4 (quad-filar wound) primaries of 58V each, wired as parallel-series ? This will mean that the secondary would only have to be specified as 29V for 8ohm and 20V for 4ohm. And the

VA rating can now be reduced to ~200VA, so that a 250VA core would be sufficient, with enough margin for input current, as each primary coil is rated for 1A, i.e. 2A total in parallel-series.

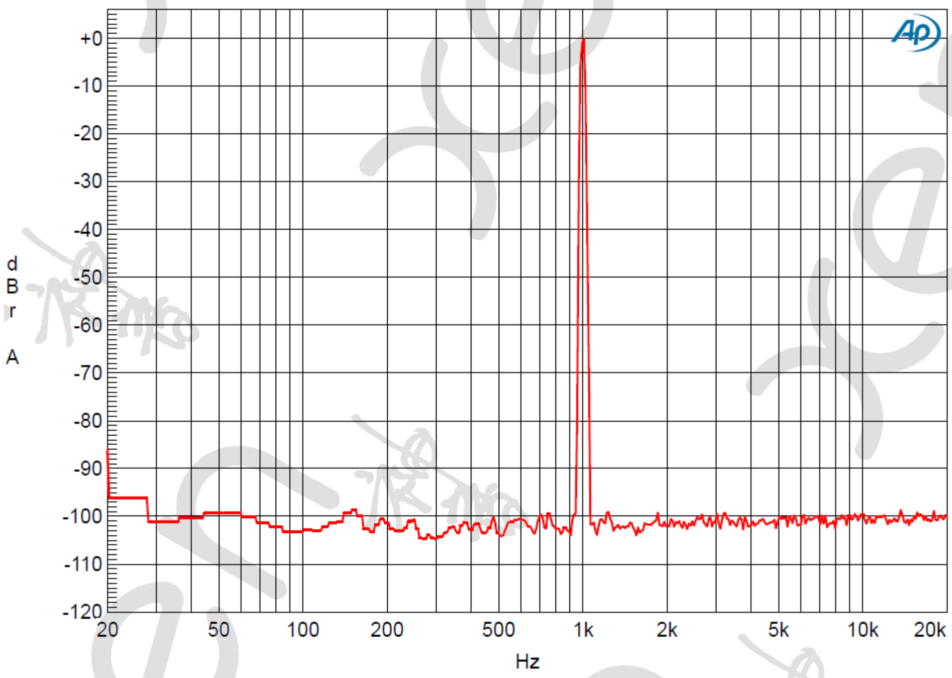
This assumes that the amplifier is used in Class A with a maximum output of +/-16V into 8 ohm. For operation in Class AB up to +/-24V (36W into 8 ohm), the maximum primary current is still below 1.2A. The limiting factor to Class AB output is the voltage headroom for the MOSFET drain, which is determined by the supply voltage as well as resistive losses in the transformer.

Testing the Transformer

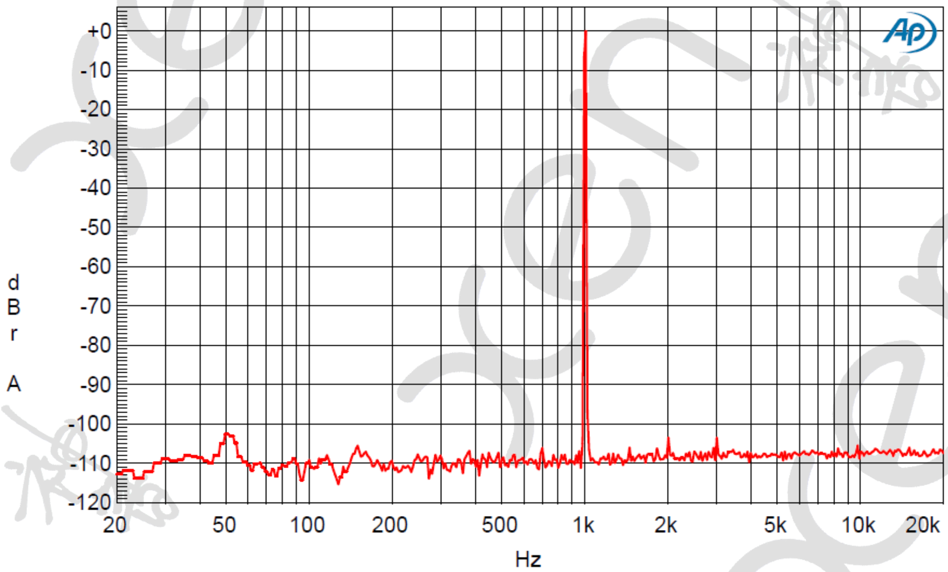
We managed to find a toroidal manufacturer willing to do such one-offs for us, got one made to measure and test. The prototype transformer was tested as follows :

1. Measure all coil inductances with LCR meter, referring to methods used to measure power transformers
2. Measure all coil DC resistances using multimeter, or calibrated DC current source and voltmeter, leaving all others open and unconnected.
3. Load the secondary with an 8R 100W power resistor. Drive the primary windings in series with a low-distortion power amp with 11.2Vrms sine wave over a range of frequencies. Measure the voltages across the 8R with a scope. Look for top and bottom end frequencies output voltage drops to 2Vrms.
4. With a low-distortion power amplifier (Purifi Class D) and a distortion analyser (AP SYS 2722), measure the distortion with the same setup as above. The prerequisite is that the distortion of the power amplifier is much lower than that of the transformer at test levels. This is a simple AC test without the DC bias current from the centre tap, and thus does not fully represent the final operating conditions. But if the bias currents are well balanced, it should not make any significant difference. And one can now test to the maximum voltage from the power amplifier before distortion jumps up near clipping.

The results were surprisingly good. The -3dB bandwidth is ~35kHz at the top end, and well below 30Hz at the bottom end.

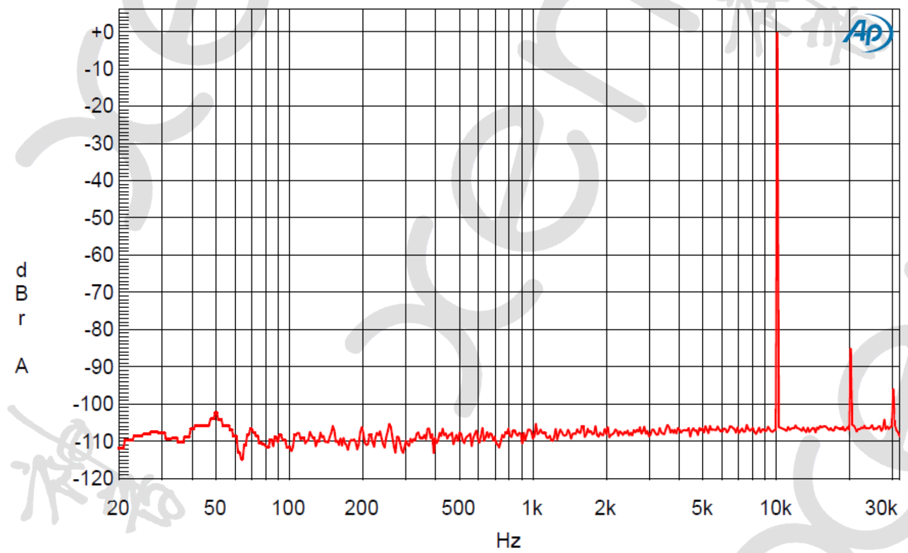


Output Transformer FFT at 1kHz 1W into 8R



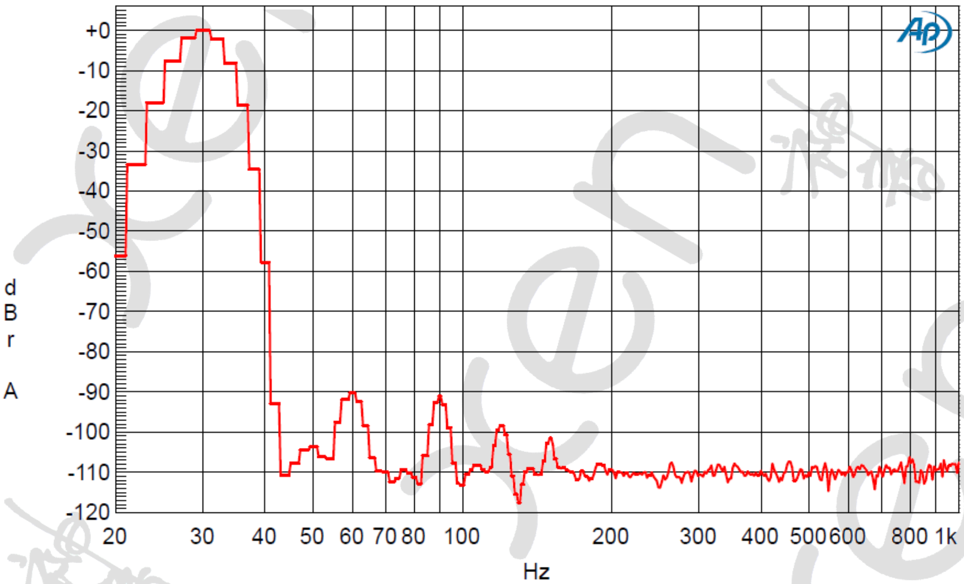
Output Transformer FFT at 1kHz 6.25W into 8R

At 1kHz 1W into 8R, distortion is below noise level. Only when the voltage is increased to 6.25W that distortions begin to show, but still well below -100dB. At 10kHz 6.25W, H2 is about -85dB, with H3 another 10dB lower.



Output Transformer FFT at 10kHz 6.25W into 8R

And at 30Hz 6.25W, both H2 and H3 are about -90dB.



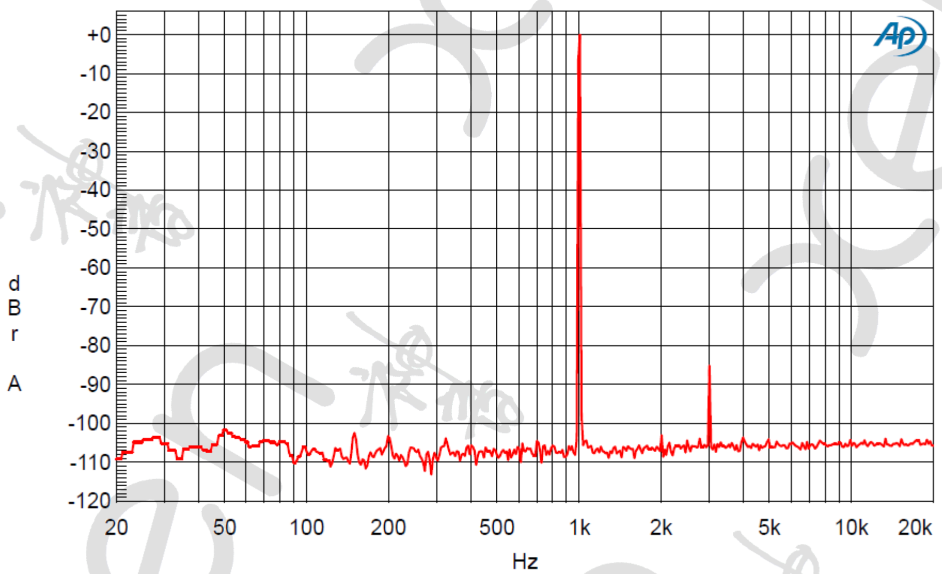
Output Transformer FFT at 30Hz 6.25W into 8R

Testing the Circuit without the Transformer

In order to be able to compare performance of different transformers, it would be useful to determine the performance of the rest of the circuit on its own. This can best be done by replacing the transformer primaries with 2x 64R power resistors at the MOSFET drains. At a bias of 420mA, these will drop an additional 27V, in addition to the original 80V. A power supply of 110V 1A would be required.

Without the transformer, the feedback network would have to be connected fully differentially from both drain resistors, and the feedback ratio converted to 40:1 to compensate for the transformer step-down. Additional coupling capacitors are also required to block off the DC of the drain resistors.

This measurement will serve as a reference as to how much the final distortion is contributed by the circuit, and how much by the transformer.



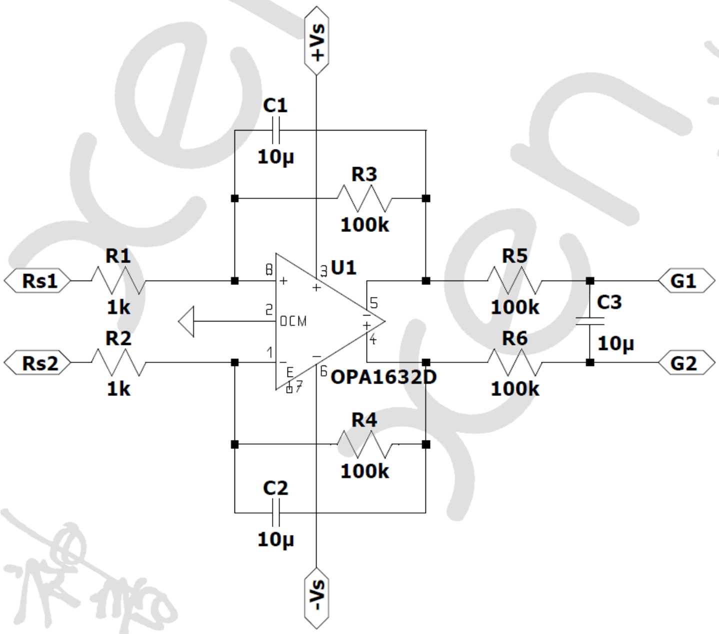
Quadrige Resistor Test, FFT at 1kHz 1W into 8R

The distortion spectrum of the Nemesis Quadrige circuit itself is shown above. The non-existence of 2nd harmonics is a proof of perfect matching, enabling even harmonics cancellations. The measurement results agree well with Spice simulations. -3dB bandwidth is > 400kHz without the feedback caps and the gate-source caps.

Bias Current Balancing Circuit

Bias control circuits are well known for tube amplifiers. One of the well-explained examples can be found at Rob Munnig-Schmidt's informative website [10]. In case of the lateral MOSFET however, things are much more favourable. Apart from little or no aging over lifetime, the BUZ900 has a negative tempco above ~100mA, so that it is self-stabilising and does not suffer from thermal run-away. Any change in bias current occurs with a long time constant, so that bias sensing can be averaged over a large time constant, for example 5~10s. Moreover, the Nemesis is designed to work largely in pure Class A (up to 16W). And as it is push-pull, we are less interested in the absolute bias value, but the difference between the two legs, so as not to pass any net DC current through the (toroidal) transformer core.

By placing a small value (< 1 ohm) current sensing resistor at the source of each MOSFET, the voltage across these two resistors can be used with a fully differential opamp (e.g. OPA1632) to generate a LPF signal to fine adjust both gate voltages. One can consider this as a differential servo for DC current. The principle of such a circuit is illustrated below :



But such a circuit is only necessary if passive stability proves to be insufficient. The source resistor can in any case be used to monitor bias with e.g. a digital voltmeter after a low-pass filter.

Some details of the circuit :

If a source resistor of e.g. 0.33R is added to both legs, the low-pass filtered value of the differential voltage across the two is a measure of bias difference. To measure 1mA bias difference, voltage difference across Rs = 330µV.

The transconductance of BUZ is about 0.8S at bias. To change 1mA bias, Vgs has to change by 1.25mV minimum. Allowing for 20dB NFB, Vgs should change by 12.5mV or more for 330µV input. So the minimum gain of the above circuit should be 38. For gain of 100x, R3,4 are to be 100k, and their value can be increased should more gain be desired. With 100x, 1mA bias difference = 33mV at differential output. The offset voltage of OPA1632 is specified at 0.5mV, which is equivalent to bias current difference of 1.5mA. The input bias current is 2µA, causing 2mV drop across R1,2. But the differential bias current is 0.1µA, equivalent to bias current difference of 0.3mA. So the circuit should be able to hold the bias current difference to 2mA or so.

It is of course important to make sure the two current sensing resistors are well matched to better than 0.1%. This can be done by passing a constant current of say 1A through the two sired in series, and measuring the voltage across them with the same voltmeter. Fine trimming can then be done by paralleling the one with the higher resistance with a suitably low-tempco metal film resistor.

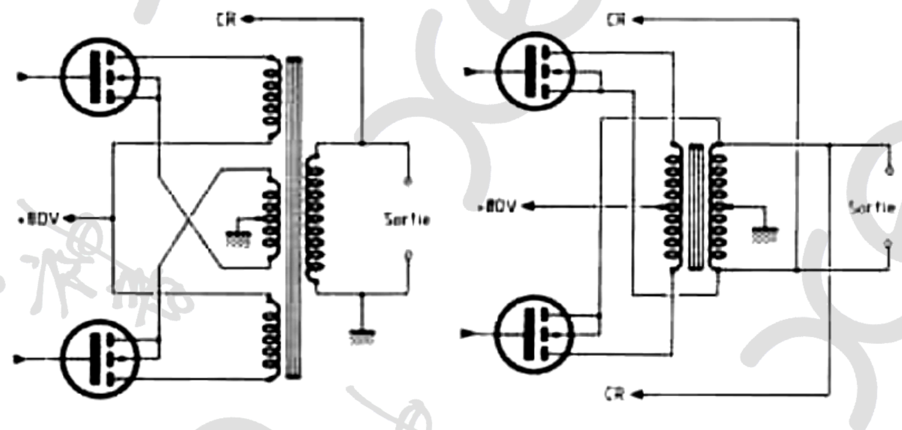
In our prototype, the voltage difference was measured across the two source resistors, which were pre-matched to better than 0.1%. The actual value of the bias current is of less interest, only the difference between the 2 legs is important. Over a period of 30 minutes, the difference between the two legs was hardly more than 3mA. The added complexity of the bias current balancing circuit is therefore not justified.

The "Booster Quadrige" with Balanced Inputs

The circuit is almost totally symmetrical, so why not use it as a fully balanced amplifier with differential inputs ?

This can easily be done by configuring the frontend diff-pair as such, but requires a transformer with an additional centre tap to be added at the output terminals, with the midpoint connected to Gnd. The output terminals are then also symmetrical relative to Gnd, and cab be used as feedback signals back to the differential front end.

In the original article of the Nemesis Quadrige, Hiraga also published a couple of alternative feedback schemes, one of which made use of this centre tapped output :



But in the following issue of the L'Audiophile, only a brief message was given as follows :

"Booster Quadrige
Many apologies to our readers: Jean Hiraga was not able to publish in this issue the rest of the article concerning the Quadrige. As indicated at the end of the last article, the Quadrige was equipped with a new transformer output matrix that turned out to be much more efficient than previous. For a 50W

output we have obtained in the absence of any negative feedback network bandwidth between 20 Hz (-0.5 dB) and 30kHz (-3dB). Applying a feedback rate of 12 dB, we obtained a substantial improvement in linearity (15Hz ~ 20kHz at -0.2dB). The JFET differential circuit turned out to be ill-suited after the application of transformer feedback, making the case to continue with a tube differential input stage."

Quick simulation of the output circuit on the right confirmed that the local transformer bootstrap at the MOSFET source does improve distortion, making it a nice unity-gain buffer without the frontend. So now, what to do with the frontend ?

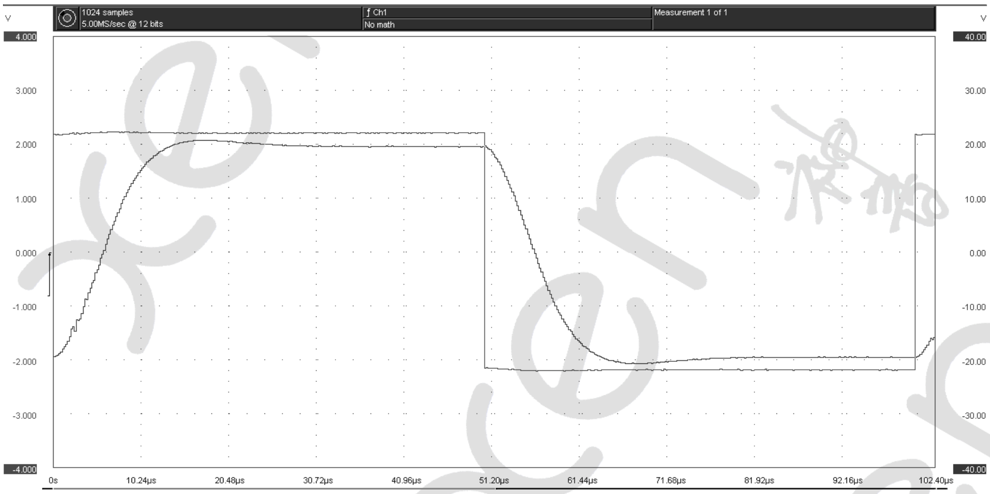
One can use the same frontend configuration as before, and apply differential feedback. But because of the bootstrapping of the MOSFET source, it has become a much more difficult load to drive. Even changing the JFET to 2SK170 or 2SK369 does not really improve the overall performance. Maybe that it was the reason why it was not continued down that route. On the other hand, the original Nemesis Quadrige proves to be already quite near the optimum. If a solution with no global feedback is desired, the Zeus (follower) mentioned at the beginning, together with a high-gain (~50x) FET or tube frontend, is perhaps a better solution. The same transformer can still be used.

Amplifier Distortion Measurements

The transformer was connected to the transistor circuit and the amplifier was measured as a whole, again with 8R resistive load. With the reactance of the transformer in the circuit, the amplifier becomes unstable without the feedback cap and the gate-source caps. These were restored, ad the amplifier was stable with a reduced bandwidth of 50kHz.

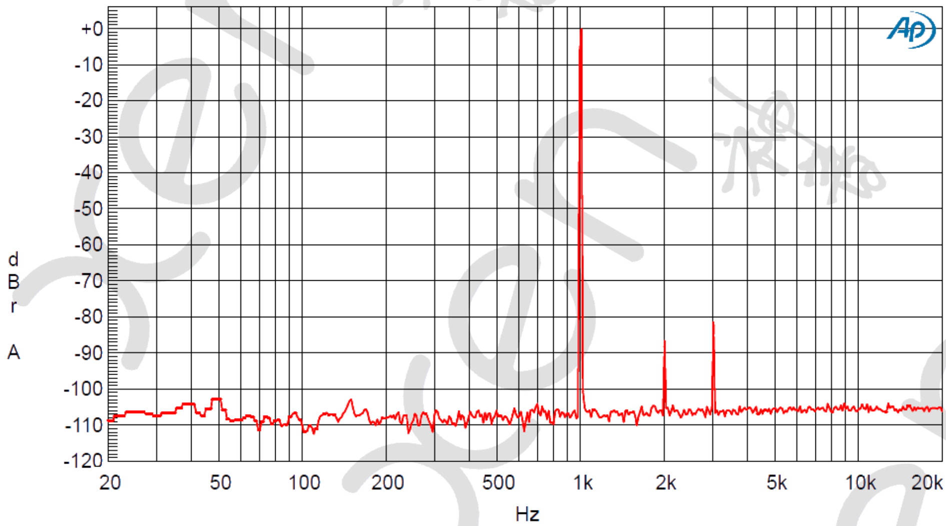
Nemesis Quadrige Frequency Response

The 10kHz square wave has a slow drop-off after the sharp rise, probably due to the LF end of the frequency response. No humps or local resonances can be seen in the frequency response curve up to 2MHz.



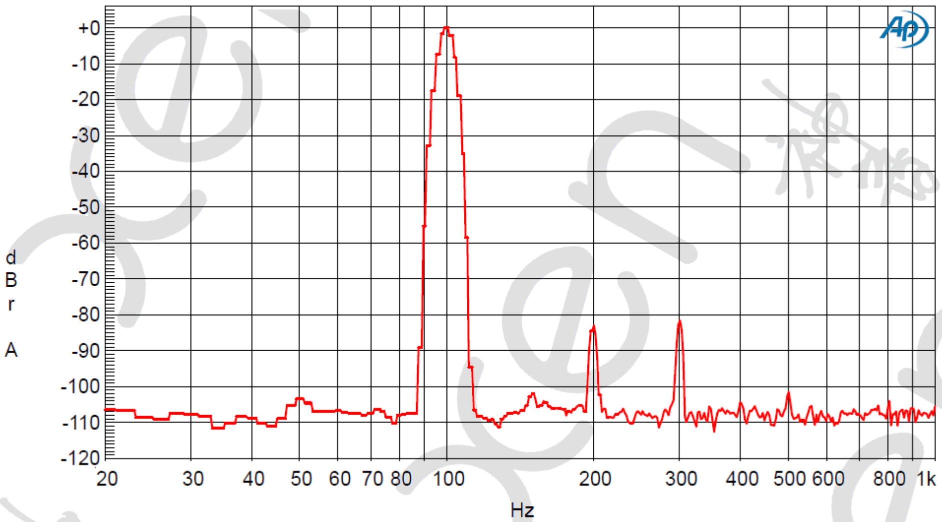
Nemesis Quadrige 10kHz Square Wave

At 1kHz 1W into 8R, the 3rd harmonics was essentially the same as that of the transistor circuit alone, but a 2nd harmonic a few dB lower now appeared. This is likely to be caused by slight asymmetry in the transformer windings.

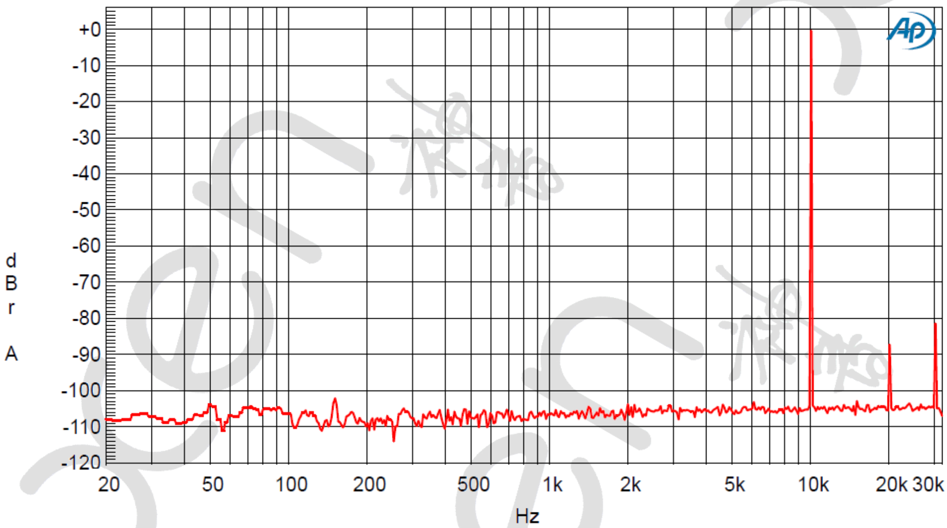


Nemesis Quadrige FFT at 1kHz 1W into 8R

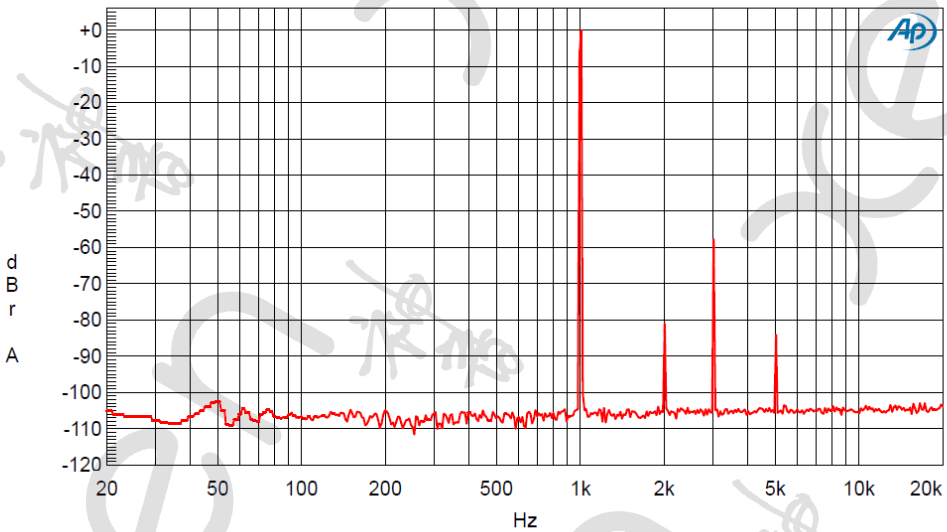
What is amazing is that the distortion spectrum is completely flat with frequency, and hardly changes when going from 100Hz to 10kHz. This can be contributed to the lateral MOSFET's low capacitances, but also its linear with frequency. The same cannot be said for other MOSFET types.



Nemesis Quadrige FFT at 100Hz 1W into 8R



Nemesis Quadrige FFT at 10kHz 1W into 8R



Nemesis Quadrige FFT at 1kHz 1W into 8R

As to be expected, distortion increases both with amplitude. At 12.5W into 8R, H3 rises to -60dB, but H2 and H4 are 20dB and 25dB lower.

So here you are….

Simple circuit with gain, few components, no expensive transformers. The Hiraga Nemesis Quadrige has been reborn. It comes as a surprise that this has not received more attention over the decades.

References

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