

The 60-Watt Quadrige Amplifier

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Choice of circuit

This had already been the subject of a pre-selection of a first project featured in issues number 34 and 35 from L'Audiophile. The push-pull circuit, the problem of single ended input made the use of minimum four transistors obligatory, two for power and two for phase shifting and add to overall gain. While the final solution adopted was that of a differential input stage, many other solutions were possible. It would for example be possible to take the first two stages of the 20 Watt class A amplifier or Le Monstre. In these two cases, we would however use a total of six transistors, four of them being reserved for the input stage. Circuit A of page 21 of Issue 34 of L'Audiophile is also attractive because it only uses 4 field effect transistors. The input transistors being complementary, the choice could have focused on 2SK147 / 2SJ 72, or 2SK170 / 2SJ74. The only precaution to take would have been to match the transistors well. Circuit B is an extrapolation of circuit A with complementary differential pairs. It is a "fashionable" input circuit today which tends to get complicated very easily. The complementary pair (in dual package) 2SK240 / 2SJ75 could have fitted perfectly in this case. A good design, on the other hand, requires the use of current sources between supply rails and JFET sources, which can bring the number of transistors used on the input stage to eight or ten.

So we arrive at Circuit C, which uses a single differential input pair. It's the one which will be retained, at least provisionally. It does not always have to be the best solution on the first try. It would also have been possible to remove the input stage and replace this one with an input transformer with centre tapped secondaries, in a setup similar to those of the first transistor amplifiers, the difference being that these were in class B, and that the transformers were of mediocre quality and that the power output hardly exceeded 1 watt. The major drawback of an input transformer would have been its high price, given the quality required: approximately 100 kHz bandwidth, and primary to secondary capacitance of nearly 600 pF (in addition to the resistive load). These 600 pF correspond to C_{gs} , the gate to source capacitance, the Gnd being connected to the source. The Power MOSFET 2SK135 offers excellent transconductance characteristic (3MHz cut-off frequency), which is limited by the input capacitance C_{gs} of 600 pF and by the intrinsic resistance of the gate, equivalent to R_{be} of bipolar transistors and which reached the value of 65 Ω in the 2SK135. A big advantage of input transformer would have been the low internal resistance of the secondary. The equivalent circuit of 2SK135 is depicted in Figure 4.

It should be realized that Y_g is placed in series at the input and its association with the C_{gs} forms a high pass filter. Taking an internal resistance of the input circuit as 0 Ω , the bandwidth would be 3 MHz (at -3dB). Now, under this condition, we realize that even in charging the MOSFET 2SK135 in source follower, we could get a theoretical bandwidth of 28 MHz (!). But it would be enough, according to the calculations given by the Hitachi engineers Sampei, Ohashi and Ochi (principal designers of the 2SK135 and derivatives) for a series inductance of 50nH at the input of the circuit to cause oscillation at a frequency between 60 and 80MHz. This would correspond to a gate connection of a few millimetres in length. The equations concerned are given in Figure 5. It follows that the preventive measures against these risks of parasitic oscillations can only be obtained, as we have seen for the Nemesis circuit, by limiting the bandwidth in these approaches:

- Increase in R_g (internal resistance of the input circuit)
- Selective drain-gate feedback reducing from a certain frequency
- Choke placed in series with the load
- Low-pass filter placed at the input
- Input-output selective feedback.

In the case of the Nemesis, the precautions taken in this respect were noted, the three anti-oscillation circuits having nevertheless made it possible to obtain more than 100 kHz of bandwidth after application of the feedback loop. As was said in Nos. 34 and 35, these compensation networks are also to be adjusted according to the wiring and especially according to the orientation of the output

transformer. Since the phase shifts at low, half or full power are only revealed once the assembly has been carried out, it is difficult to dissociate the origins of these oscillations. It would no doubt be possible to carry out bandwidth and phase measurements on the 2SK135 alone by taking pure resistors as a load (heatsinked models), then to measure the output transformer alone, by applying to its primary the signal from a power functions generator as well as a DC current of value close to the real current crossing the primary. We notice, even on very high-quality output transformers, that if the linearity is perfect between a few hertz and 100 kHz, significant phase shifts and sorts of sawtooth on the frequency response occur at higher frequencies, between 120 and 250 kHz (in general), phenomena always disturbing even if they occur at 20 or 30 dB below the linear range. However, the frequencies indicated above relate to an output transformer for using with tubes, the primary of which would have a value between 3 and 6k Ω . On 64 or 128 Ω , and by building a very high-quality transformer, these parasitic phenomena could occur at much higher frequencies. The goal is therefore to seek a good association of elements limiting the bandwidth more or less naturally (input stage, wiring, output transformer) to the super-fast MOSFET (switching time 20x faster than a conventional power bipolar transistor, like 2N3055) and with a very high cut-off frequency.

Going back to the input stage, it must have a relatively low output impedance for another reason than the one mentioned above. Contrary to low power tubes on which the practically zero grid current allows a pure voltage drive or almost, a MOSFET of the 2SK135 type requires a voltage drive (or almost) below 5 and 6 kHz, which gradually turns into power drive as the frequency increases. At 250kHz and on a 100-watt OTL push-pull assembly, it would have taken 100mW to properly drive the gate of the 2SK135, while 0.1mW would have been enough if it had been a question of a frequency of 750Hz.

This power drive versus frequency characteristic is shown in Figure 6. This is why a value of 10 k Ω was used on the Nemesis input. Which may seem a bit low but was desirable. On the Quadrige, this problem is solved by the insertion of the differential input stage with JFETs, which makes it possible to bring the value of the input impedance to 100k Ω . We can, on the other hand, compensate for any loss of sensitivity (if a C.R. loop is applied) thanks to the gain provided by this first "amplifier-phase-shifter-driver" stage. This solution is, in terms of practical use, much more universal than that of an input transformer whose primary should have been lowered to 200 or 600 Ω .

The Quadrige therefore uses a differential input stage made up of two field-effect transistors. A preliminary test concerning the 2SK240 transistors had been made because of the remarkable matching of the two distinct transistors placed inside a unique metal box. Too little gate voltage (V_{th}) offered too narrow working range, leading too quickly to pinch-off (or cut-off) or into the positive bias region of the gate. The final schematics is similar to that described in L'Audiophile issue 35, Figure 10 (page 15). The two 2SK30AGR transistors are therefore mounted as a differential amplifier, with drain loads of 4.7k Ω . This previously chosen value later proved to be a good compromise during the experiments that followed. The source resistance R_s ; of value 240 Ω , made it possible to obtain the desired polarization value, i.e. -0.6 V. Under this condition, we have unfortunately observed a serious imbalance in the AC output voltages V_{D1} and V_{D2} , this being due, as Gérard Chrétien pointed out, to the fact that a good balance of the output voltages can only be obtained by taking values of R_D (drain load resistors) much smaller than the value of R_s ; the common source resistance.

In practice, bipolar power supplies solve this question, so to speak, automatically.

$V_{D1} = -V_{D2}$ in the ideal case of a perfect differential (R_s / R_d to get as close as possible to a current source on the JFET sources).

The negative reference being grounded, we can therefore see that it is necessary to place the tail of the common source resistor R_s , the value of which will be at least twice that of the drain, at a fairly large negative value with respect to ground. For practical reasons, approximately -40 V was chosen, which leads to a value of R_s of the order of 18k Ω . This new condition, slightly complicating the power supply circuit, provided very good measurement results: perfect symmetry, impeccable phase shift as well as amplitude / frequency bandwidth exceeding 100kHz. The V_{DS} voltage has, on the other hand, been increased, the range 10~18V being more favourable: a little less distortion, output signal of greater maximum amplitude. About the rate of distortion, a curious point is to note that we can obtain

at the output of the amplifier a distortion rate noticeably lower than that produced by each of the two stages taken separately. for which the overall distortion rate is only one third of the distortion produced by the first or second stage taken separately. It is easy to understand that this is the distortion cancellation method, obtained by subtracting two distinct distortions. The crushing effect of a sinusoid coming on the first stage from working in the compressed zone of the curves, located closely to the cut-off can "expand" again to resume its original shape, or almost, on the second stage which then receives a signal out of phase with respect to the input, which will be made to operate in the more expanding zone of the curves. A judicious calculation, sometimes even obtained by graphic calculation, can make it possible to obtain good results, the main risks being however the possibility of generating an irregular harmonic distortion spectrum according to the amplitude and the frequency. Let's cite the case of a 12 BH7 + 300 B single-stage tube amplifier which, according to the "Stereo Gallery" formula (name of the device marketed in Japan in 1969) produced only 0.15% distortion at 6 watts, without using feedback. Closely examined, this amplifier revealed in fact a nonlinear distortion/power characteristic and also the appearance of harmonic distortion in the 5th and 7th orders. This is why it was considered preferable, at first, not to enter into this game.

The input stage, modified version

Figure 7 shows the small modifications made to the input stage, compared to the first draft.

Note that the Zener diode goes from 13 V to 22 V and that the series supply resistance goes to 4.7k Ω (instead of 8.2k Ω). The coupling capacitors have the value of 2.2 μ F. We also note the presence of a small phase correction network (R-C series) placed between the two drains of the 2SK30AGR. Note the absence of series resistance at the input (normally intended to limit the bandwidth at high frequencies). The input is, on the other hand, direct-coupled without any input capacitor.

It should be added that various experiments are still in progress on all the circuits of the Quadrige.

The output stage

There were some hesitations about the push-pull output stage, about power and operation class. In Class A and boosting the power supply to 75 V, it would have been possible to obtain around 35 watts. Apart from class A, four other operating classes could have been chosen :

- Class B, excluded in this case, due to the risks of its non-linearities and the generation of crossover distortion. Class B would on the other hand have required driving the gates of the MOSFETs with a lower impedance.
- Class AB, for which the operating point moves over practically the entire load line, including in the slightly positive region of the gate, the quiescent current being reduced to a low value.
- Class A2 which is very close to Classes A or A1; for which the quiescent current is a little higher than in class A, the operating point being able to move into the slightly positive region of the gate.
- Finally, Class AB1 which combines the advantages of high power and low distortion in a push-pull assembly. The operating point moves along the entire load line, but does not enter the positive region of the gate. The quiescent current is a little lower than in Class A, which avoids the creation of crossover distortion.

Experiments made from a separate adjustable power supply have achieved just over 60 watts output, with guaranteed true Class A operation up to about 17 watts (calculated value). The gate bias goes to +1.54V (instead of +2.4V for the Nemesis) and the quiescent current goes to 1.4A for the two transistors, or 700 mA per transistor. It should be noted in this connection that the crossover distortion only appears for a quiescent current of less than 50 mA. With 1.4A of quiescent current for the two transistors, this risk is overcome. On the original project, another solution had been imagined and considered, according to the classic method of Class change A \rightarrow AB, in simultaneously modifying the supply voltage and the bias current from 42V Rail & +2.2V Vgs (Class A) to 80V & +1.5V (Class AB). This operation in Class AB, with a current slightly higher than normal, ensures operation in Class A up

to more than 15 watts and in a still linear part of the curves, with a gradual transition to Class AB, for higher powers.

The output transformer

Additional details will be given later in this subject, several experiments on prototypes being currently in progress. It is quite possible that the circuit evolves towards a symmetrical structure of the output transformer, with cross-feedback of the sources, according to a known method that could be found on old or recent assemblies, with tubes or transistors made by Quad , Audio Research or Luxman. Figure 8 illustrates the two possibilities that are currently being tested and measured.

Performance

These largely depending on the qualities of the output transformer, it is not yet possible to provide readers with definitive figures, since the prototype transformer used was not designed to work for over a dozen of watts. In class AB, it was however possible to obtain, with only 6 dB feedback, a bandwidth of between 20Hz and 75kHz, to within 3 dB, a distortion rate regularly rising with the increase in output power, ranging from 0.01% (0.1 W) to about 2% (60 watts). But these are only provisional performances. The push-pull version of the Nemesis is therefore very promising and there is no doubt that it will lead to a final version of a simplicity still unknown to date, in terms of power performance, bandwidth and distortion worthy of current hi-fi standards. The preliminary results and measurements concerning the sinusoidal and square waveforms are given in figure 9. It can be seen that these are very promising, and that if very good results have already been obtained, they deserve an optimization of the circuits and of the output transformer with a view to obtaining an even more thorough result.

Assembly

The assembly of the Quadrigé will be described in detail in the next issue of L'Audiophile. In the meantime, the two other versions described in Figure 8, comprising the cross-source feedback loops, will be tested and compared with the basic version described in this issue. It will of course be a question not only of the measurements but also of the listening results. Another test, currently in progress, will concern the replacement of the differential input stage with field effect transistors by a double triode tube. Trials are in progress at the time of going to press. They relate to the choice of the tube, the insertion or not of a transistorized current source in the cathode circuits, as well as the possibility of setting up a high voltage hybrid regulated power supply, the advantages or shortcomings provided by this, as well as the final choice. The wiring of the output transistors having to include very short links. There will also be a question of the heatsink, its shape and its optimal dimensions. It seems on the other hand it is quite possible to mount the Quadrigé in a chassis of the same dimensions as that of the Nemesis.

Addendum — Booster Quadrigé

L'Audiophile Issue No. 37, Winter 1985.

Many apologies to our readers: Jean Hiraga was not able to publish in this issue the rest of the article concerning the Quadrigé. As indicated at the end of the last article, the Quadrigé 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 (-0.5dB) 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.

Quadriga Amplifier

L'Audiophile Issue No. 39, Autumn 1986.

It's the L'Arlésienne, you might say. Nothing yet in this issue concerning the final version of the Quadriga. It's not for lack of being questioned all week long about this project. Some comments are in order. As you know, various experiments were carried out with several prototype output transformers. Nevertheless, inherent problems with the transformer remain and cause us not to be completely enthusiastic. In addition, the numerous tests, both in terms of measurement and listening, carried out over nearly six months on the output transformers for single-stage assemblies have been extremely instructive. As can be clearly noted, the existing differences between models of similar specifications are colossal (ref. article in this issue). Resorting to compensations, corrections, artifices of various kinds is not very exciting, the final performance depends directly on it. Moreover, the possibilities of collaboration with French manufacturers remain terribly limited, which inevitably slows down the evolution of this project. In addition, it is a very special technology, which does not help matters; obviously, if we had manufacturers such as Tango or Partridge "on hand", the problems would be solved much more quickly. We take this opportunity to appeal to MM. French manufacturers of output transformers wishing to explore the issue with us...

Figures

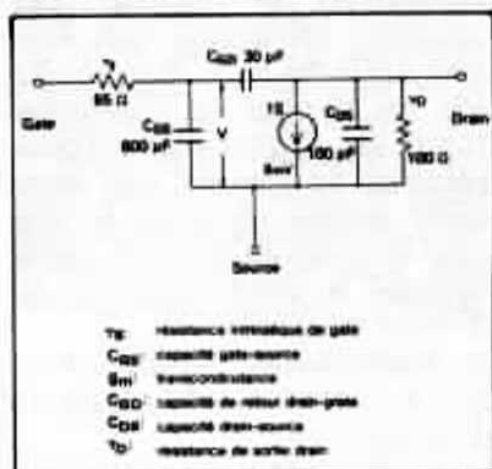


Fig. 4 : Schéma équivalent du transistor MOS-FET 2SK 135.

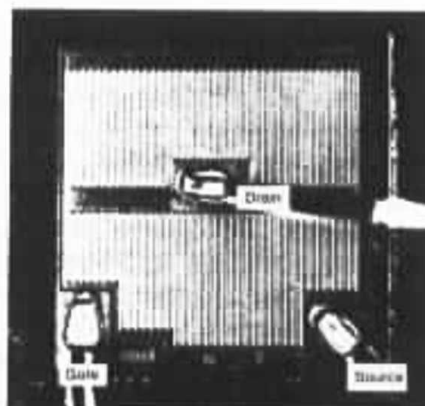


Photo du substrat (puce) du MOS-FET 2SK 135. La puce mesure $4,5 \times 4,5$ mm. Le canal N a une longueur totale de 40 μm et une largeur de 9 μm . (Document Hitachi)

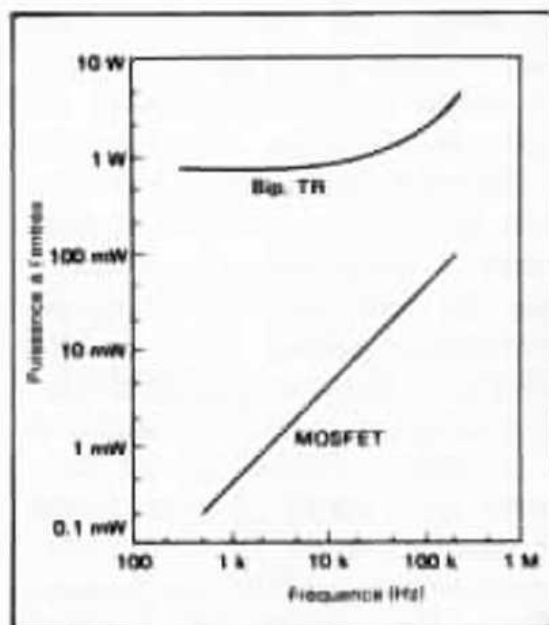
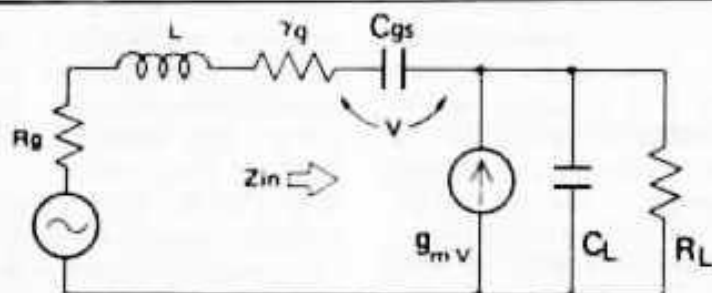


Fig. 6 : Caractéristique d'attaque en puissance de l'entrée des MOS-FET (pour un amplificateur de 100 watts) en fonction de la fréquence. Comparaison avec celle d'un transistor bipolaire.



L'impédance d'entrée est donnée par
la formule :

$$\begin{aligned}
 Z_{in} &= \gamma g + sL + \frac{1}{sC_{gs}} + \frac{R_L}{1 + sC_L R_L} \left(1 + \frac{g_m}{sC_{gs}}\right) \\
 &= \gamma g + sL + \frac{1}{sC_{gs}} - \frac{sC_L R_L^2}{1 + \omega^2 C_L^2 R_L^2} \\
 &\quad - \frac{s g_m R_L}{(1 + \omega^2 C_L^2 R_L^2) \omega^2 C_{gs}} + \frac{R_L}{1 + \omega^2 C_L^2 R_L^2} \\
 &\quad - \frac{C_L R_L^2 g_m}{(H \omega^2 C_L^2 R_L^2) C_{gs}} \dots \dots \dots (2)
 \end{aligned}$$

à condition que :

$$\begin{aligned}
 \gamma g + \frac{R_L}{1 + \omega^2 C_L^2 R_L^2} - \\
 \frac{C_L R_L^2 g_m}{(1 + \omega^2 C_L^2 R_L^2) C_{gs}} < 0 \dots \dots \dots (3)
 \end{aligned}$$

Lorsque l'impédance d'entrée Z_{in} possède une
partie réellement négative. Si la somme de
 $R_g + Z_{in}$ réel est négative, le circuit entre
en oscillation. La fréquence d'oscillation
est donnée par la formule :

$$\begin{aligned}
 sL + \frac{1}{sC_{gs}} - \frac{sC_L R_L^2}{1 + \omega^2 C_L^2 R_L^2} \\
 - \frac{s g_m R_L}{(1 + \omega^2 C_L^2 R_L^2) \omega^2 C_{gs}} = 0 \dots \dots (4)
 \end{aligned}$$

$$1 > \omega^2 C_L^2 R_L^2, \quad 1 > \omega^2 C_L C_{gs} R_L^2$$

$$f_{osc} \approx \frac{1}{2\pi} \sqrt{\frac{1 + g_m R_L}{L C_{gs}}} \dots \dots \dots (5)$$

Fig. 5 : Calcul de l'impédance d'entrée et des conditions pour lesquelles le circuit devient instable.

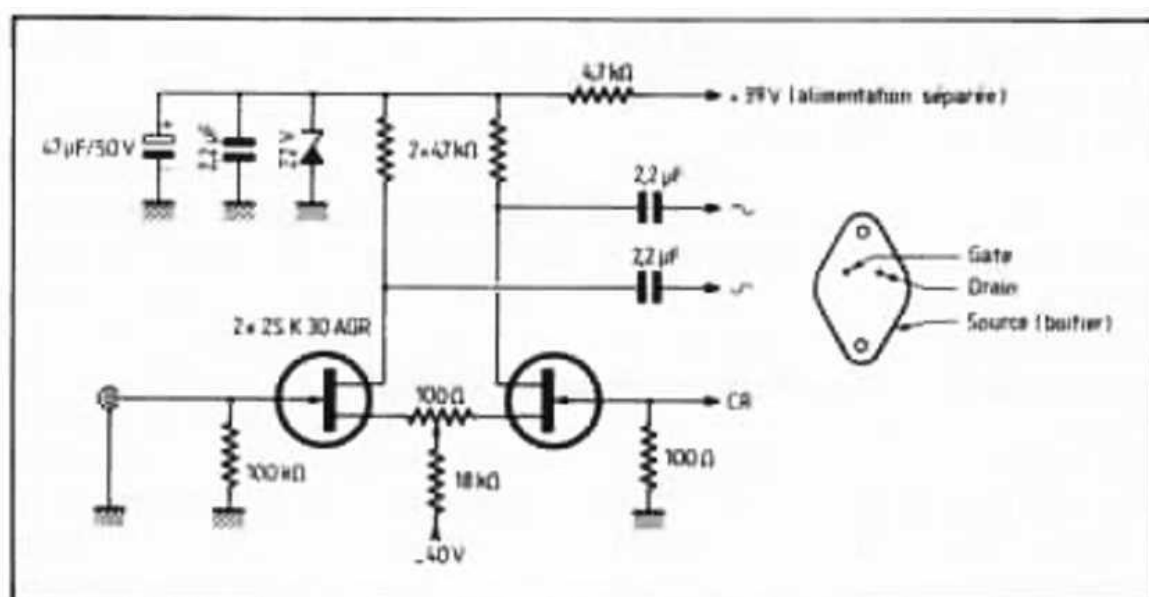


Fig. 7 : Etage d'entrée, après modification.

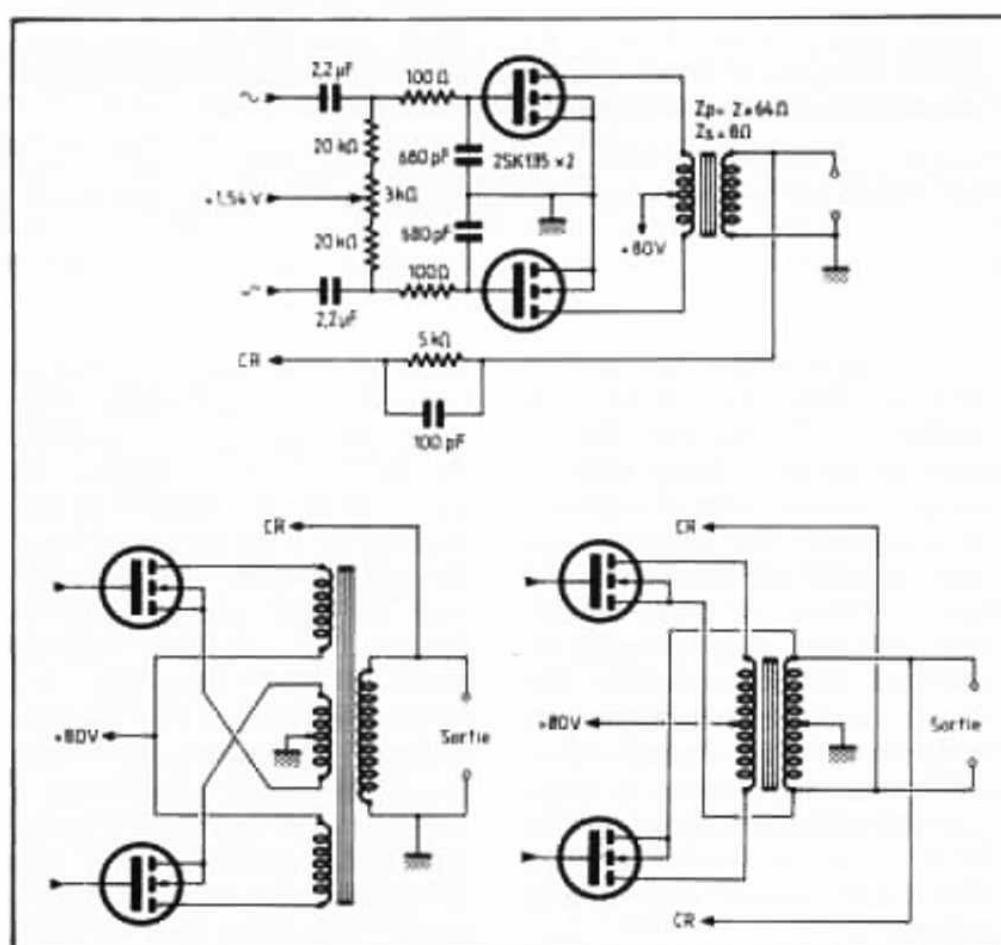
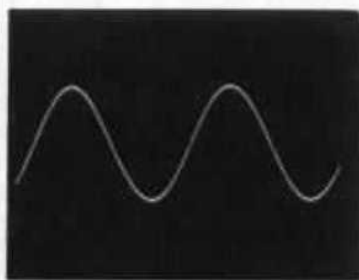
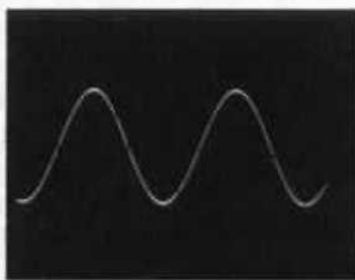


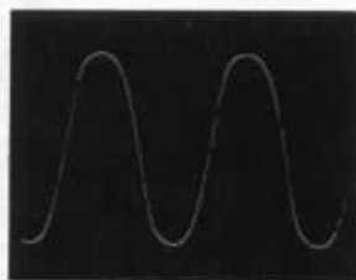
Fig. 8 : Etage de sortie du Quadriga et deux autres projets concernant l'application de deux boucles de contre-réaction de source croisées.



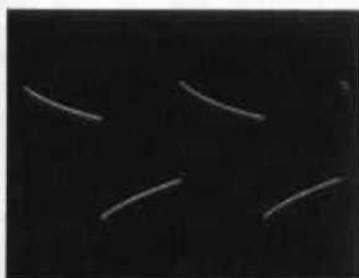
Forme d'une sinusoïde de 100 kHz sous 0,5 watt. Remarquer l'absence de distorsion.



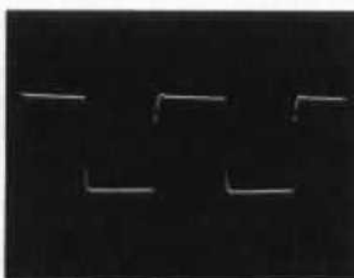
Réponse à 1 MHz, 0,05 watt. Noter l'absence de déformation de la sinusoïde.



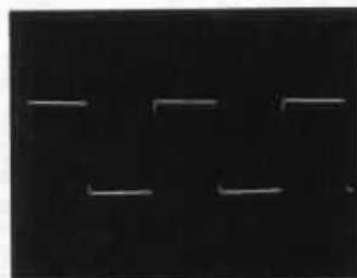
Forme de la saturation d'une sinusoïde (5 kHz) à 60 watts. Remarquer la forme arrondie et sans accident de la saturation.



Signal carré à 50 Hz, 5 watts. La réponse est excellente et on ne remarque ni dépassement ni suroscillation.



Signal carré, 2,5 kHz avant correction. Noter l'irrégularité présente sur la montée du signal.



Forme du signal carré, 2,5 kHz, obtenu après correction (680 pF sur les gates des 2SK 135 et 100 pF sur le réseau de C.R.).

Fig. 9 : Forme des signaux obtenus sur le Quadrigé, dans la version décrite dans cet article.