10.4 Phase-Splitter

A single power tube (class-A operation) allows only for small output power. High power needs push-pull operation (Chapter 10.5). A push-pull output stage requires two drive signals shifted by 180° relative to each other. These two anti-phase signals are generated in the so-called phase-splitter circuit using one or two tubes. In essence, there are three circuit-concepts: the tube operating with v = –1 in common-cathode configuration (paraphase-circuit), the cathodyne circuit, and the differential amplifier in common-grid configuration.

10.4.1 Common-cathode circuit (paraphase)

This is a simple concept: one triode provides amplification with its plate-voltage serving both as drive-signal for one of the two output tubes, and – attenuated via resistors – as drive signal for the other triode. The latter feeds its (opposite-phase) plate-voltage to the other power tube (**Fig. 10.4.1**).).

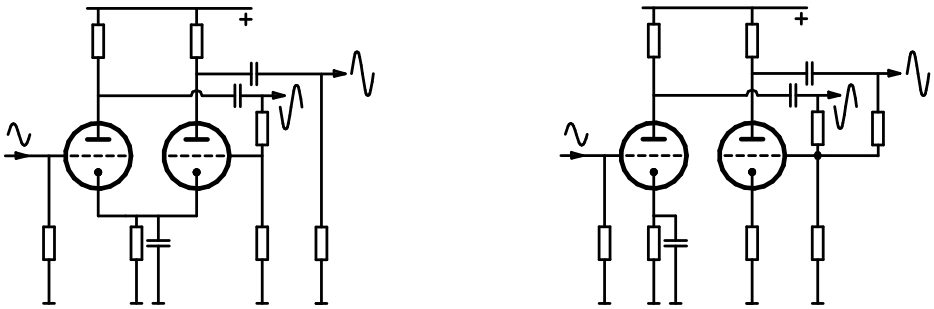


Fig. 10.4.1: Phase-inverter in common-cathode configuration. Right: modified version with negative feedback.

This basic **paraphase circuit** is predominantly found in early guitar amplifiers (e.g. the 1947 Fender Deluxe). It was soon first modified and then replaced by the cathodyne circuit. The advantage of the paraphase circuit lies in its high voltage gain and the relatively large output voltage swing of the two tubes. Disadvantageous is that the magnitudes of the output voltages are not exactly equal but depend significantly on the individual tube data. Matching the divider resistors leads to an individual symmetry, but this would have to be checked and re-checked as the tube ages. Of course, it is an entirely different question whether a guitar amplifier actually sounds best with complete symmetry of the output stage – however even if a lack of symmetry would be desired, this would have to be specific and not subject to random tube-variance.

The typical paraphase circuit – as it is found e.g. in the old **Fender Deluxe** (5B3) – attenuates the output AC-voltage of the first tube with a 250-kΩ/7.0-kΩ-divider by a factor of 1/44. For a precise calculation, the internal impedance of the first triode must be added in – this is approximately 50 kΩ. The second triode amplifies this attenuated voltage by a factor of -44, making available two AC-voltages of equal amplitude and opposite phase that drive the output tubes. That would be the ideal case, anyway – in reality, however, the gain of the second tube has significant scatter.

If the voltage gain of the second tube is not at its nominal value but e.g. too small by 20%, the two half-waves generated by the power amp also differ by 20%. The consequence is that this effect alone is cause for **harmonic distortion** of 4%. One may feel good or bad about such asymmetry – at Fender, it was not liked. The voltage divider at the grid of the second triode was replaced by a current/voltage **negative feedback**: the plate-voltage is tapped (via 270 kΩ) and generates an additional current in the grid-circuit. **Fig. 10.4.2** depicts the circuit of the Fender Deluxe 5D3; it is also found on other Fender amps of the same era (Super Amp 5D4, Pro Amp 5D5, Twin 5D8).

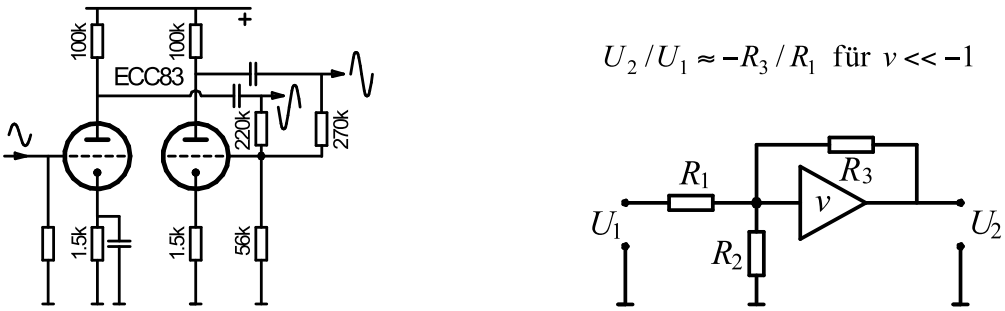


Fig. 10.4.2: Paraphase-circuit with current/voltage negative-feedback (Fender Deluxe 5D3, 1954).

The principle of the current/voltage negative-feedback is also used in the inverting OP (right-hand section of the figure): for an OP-gain approaching infinity, the voltage across R_2 becomes close to zero; U_2/U_1 is merely defined by the relationship of the resistances and not by the gain anymore [e.g. Tietze/Schenk]. For a tube circuit, this simplification holds only approximately – but the basic operation is the same: if the open-loop gain of the second triode changes by 10%, the ratio of the two (opposite-phase) output voltages changes by merely 1% due to the negative feedback. The latter stabilizes the ratio U_2/U_1 of the two output voltages – the circuit is termed "self-balancing paraphase circuit".

The negative feedback has a further effect: it reduces the **internal impedance** of the right-hand triode. With a load, the plate-AC-voltage of the triode on the right becomes smaller and consequently the voltage fed back via the 270-kΩ-resistor decreases also, resulting in a overall larger voltage gain. To some extent at least, the load-dependent decrease in the plate-voltage is compensated. The internal impedance of the triode-circuit on the left (Fig. 10.4.2) is simply the parallel connection of the internal impedance of the tube (e.g. 63 kΩ) and the plate resistor (e.g. 100 kΩ) – i.e. about 39 kΩ in our example. Considering the load (about 220 kΩ), as well, brings us to $R_{i1} \approx 33$ kΩ for the overall circuit. For the right-hand tube, the calculation yields $R_{i2} \approx 12$ kΩ (including load). The negative feedback has therefore reduced the internal impedance of the second triode-system to about 1/3rd. As long as the loading of the two paraphase outputs is negligible, the differing internal impedances do not play any role. However, the input capacitances of the power tubes and the occurrence of grid-currents can lead to load situations that cause considerable asymmetries.

Furthermore, it is necessary to consider that the input signal to one output tubes passes *one* RC-high-pass, while the input signal to the other output tube passes though *two* such filters, causing phase shifts in the low-frequency range. Similar effects happen at high frequencies: the detour via the second triode-system acts as an additional low-pass that causes phase shifts in the high-frequency range.

Fig. 10.4.3 shows the output voltages of a paraphase circuit having no negative feedback. For small drive levels, we indeed get two phase-opposed voltages of approximately equal amplitude. With increasing drive levels, triode-clipping starts to become visible – this shifts the operating point across the coupling capacitor. In the lower line of the figure, we see power-tube grid-currents (occurring from about +20 V) that limit the voltage-curves in the direction of positive values. Because the signal of the second triode is derived from the clipped plate-voltage, the second output signal is limited towards negative values, as well. The overdrive of the output tubes consequently is asymmetrical.

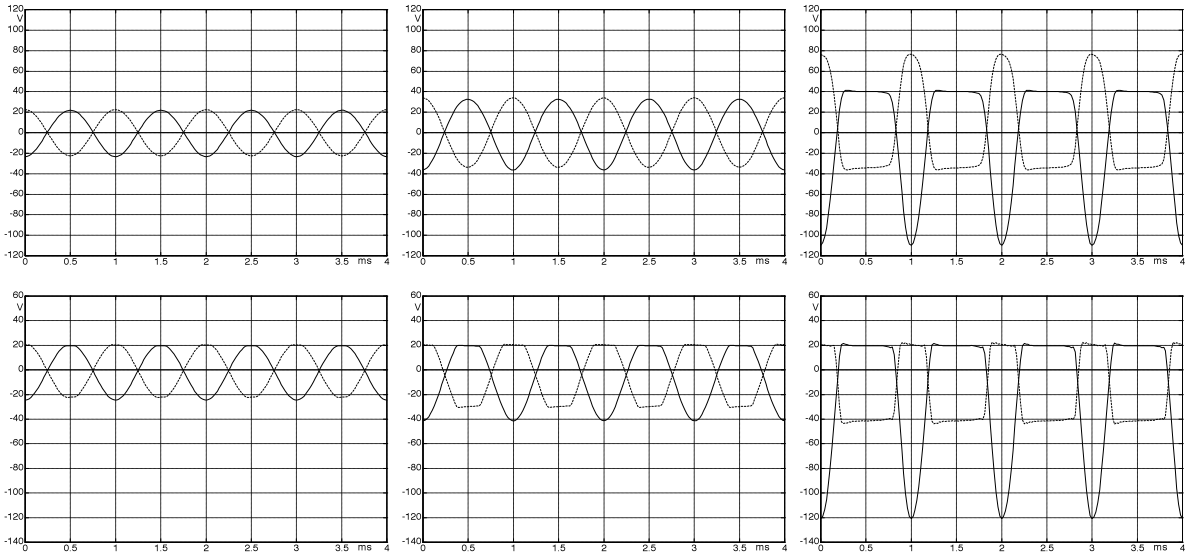


Fig. 10.4.3: Measurements on a paraphase-stage **without** negative feedback: 1st tube (——), 2nd tube (---).
Top: no grid-current limiting. Bottom; grid-current happening from 20 V. Supply-voltage for the triodes: 260 V.

Fig. 10.4.4 represents the corresponding measurements of a paraphase stage with negative feedback. We again see the different drive situations of the two power-tubes in non-linear operation. Also, the change in the duty-factor already recognizable in Fig. 10.4.3 reappears.

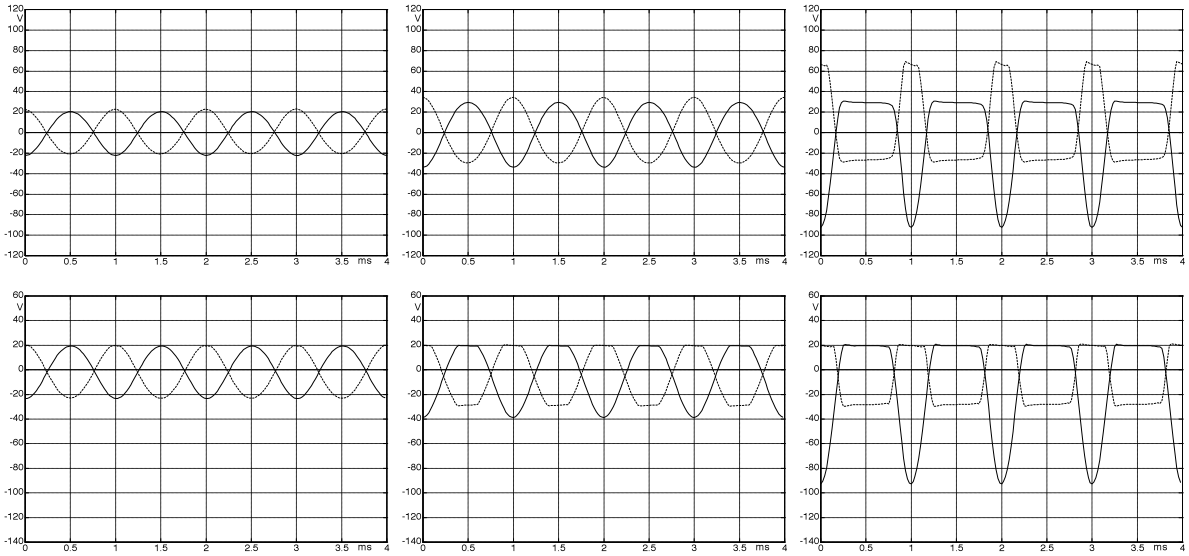
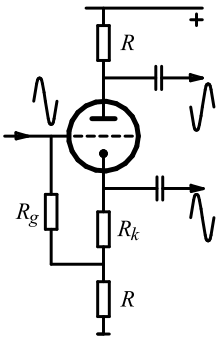


Fig. 10.4.4: Measurements on a paraphase stage **with** negative feedback: 1st tube (——), 2nd Tube (---).
Top: no grid-current limiting. Bottom; grid-current happening from 20 V. Supply-voltage for the triodes: 235 V.

10.4.2 Cathodyne-circuit (split-load)

The cathodyne circuit takes advantage of the opposite-phase-situation of the AC-voltages at cathode- and anode. Assuming a drive situation with a grid-current of zero, the cathode-current is equal to the plate-current, and therefore voltages across equal cathode- and plate-resistors will also be of the exact same amount – irrespective of any tube variances. Textbooks on circuit design tend to explain the cathodyne configuration by separating the plate-resistance into two "exactly" equal halves that then result in the new plate-resistance and cathode-resistance, respectively. It is possible that this approach led to the designers using high-precision resistors in the cathodyne-stage. For example, the schematic for the Ampeg B-42-X specifies: *all resistors 10%* – however, the caption of the 47-kΩ-cathodyne-resistors and the subsequent 100-kΩ-load resistors reads *5%*. There were even amplifiers requiring a resistor-tolerance as low as 2% for this circuit.



$$v_A = -\frac{v_k}{1 + R_k / R} \approx -v_K$$

$$v_K = \left(1 + \frac{2R + R_i + R_k}{(R + R_k) \cdot R_i \cdot S}\right)^{-1} \approx \frac{\mu}{\mu + 3}$$

$$R_E \approx \frac{R_g}{3/\mu + R_k / R}; \quad R_{iA} \approx R; \quad R_{iK} \approx \frac{R + R_i}{1 + \mu}$$

Fig. 10.4.5: Cathodyne-circuit. Signals taken directly from the cathode as is typical for Fender.

In **Abb. 10.4.5** we see a guitar-amplifier-typical cathodyne-circuit. In Fender amps, both load resistors (R) normally have a value of 56 kΩ with R_k = 1.5 kΩ and a grid-resistor of 1 MΩ. Several Fender amps received this circuit in 1955 (Deluxe, Super, Pro, Bassman, Twin) but it was only about two years until the arrival of the differential amplifier (more in chapter 10.4.3). The grid-resistor R_g of the circuit in Fig. 10.4.5 is connected to the split cathode-resistor rather than to ground. This negative-feedback arrangement substantially increases the **input impedance** R_E (in the example to about 18 MΩ). It is questionable whether the designer at Fender was aware: the coupling capacitor feeding the grid is, after all, 20 nF, just as customary with 1-MΩ-inputs. The 1-MΩ-resistor is, however, not connected to ground but to an almost equally big coherent AC voltage, and thus the effective input impedance increases (bootstrap). The 10 nF and 18 MΩ component values results in a **high-pass** cutoff-frequency of 0,4 Hz – quite generous for a guitar amplifier. Gibson used, in their GA-19-RVT, a capacitor of merely 500 pF for the cathodyne input capacitor – maybe they knew more?

The **voltage gain** from grid to cathode is about $1 - 3/\mu$, with μ = open-loop gain of the tube. For the ECC83 follows, with good approximation: $v_K = 0.97$. As is typical for Fender, the amount of the plate-AC-voltage is slightly less, about $v_A = -0.945$. The **internal impedances** of both outputs are, however, highly different: at the plate we have (with good approximation) 56 kΩ (negative current-feedback at the cathode), while no more than about 1.2 kΩ are present at the cathode (cathode-follower). Amplifier tubes are often said to present *no load* to the preceding circuits, and if that were always correct, the differences between the internal impedances would be irrelevant. However, grid-currents may flow in the power tubes, and if that is the case, plate- and cathode-voltages in the cathodyne stage start to be different.

An AC-relevant plate- or cathode-load has different effects on the respective other electrode: a *cathode-loading* would increase the plate-current and thus grid-to-plate gain, while a *plate-loading* would decrease this gain. Both types of loading would however have only little impact on the grid-to-cathode gain (negative feedback). The cathodyne-stage does experience loading by the output tubes. The latter are showing a high input-impedance only as long as the power-tube grid is sufficiently negative relative to the power-tube cathode. At full drive levels, and in particular in a state of overdrive, grid-currents do flow, and the cathodyne stage operates with a non-linear load.

Fig. 10.4.6 shows the time-functions of the plate- and the cathode-voltages for different drive-levels – first without the loading effect the output tubes have. Compared to the paraphase-circuit, the maximum voltages are smaller but the symmetry is better. As we include the loading by the power tubes (6V6, **Fig. 10.4.7**), the shape of the plate-voltage changes due to the grid-current-drain via the cathode – this increases the plate-current and consequently the voltage drop across the plate-resistor. In the cathode-voltage, there is practically no corresponding protrusion because the voltage gain of the cathode-follower is only marginally influenced by the plate-resistance. A typical effect found in tube amplifiers is shown in the last line of the figure: the supply-voltage decreases with increasing overdrive ("sagging"). Therefore, the minimum voltage is not constant but depends on the filter-circuit in the power-supply.

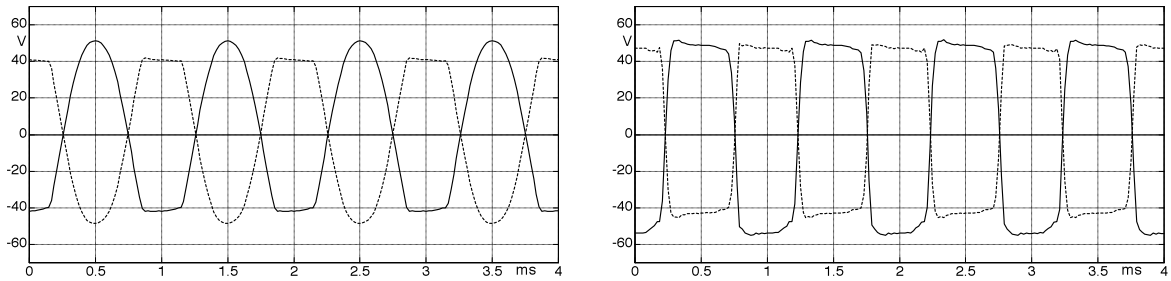


Fig. 10.4.6: Cathodyne-stage without load; AC-component. Plate-voltage (----), cathode-voltage (——).

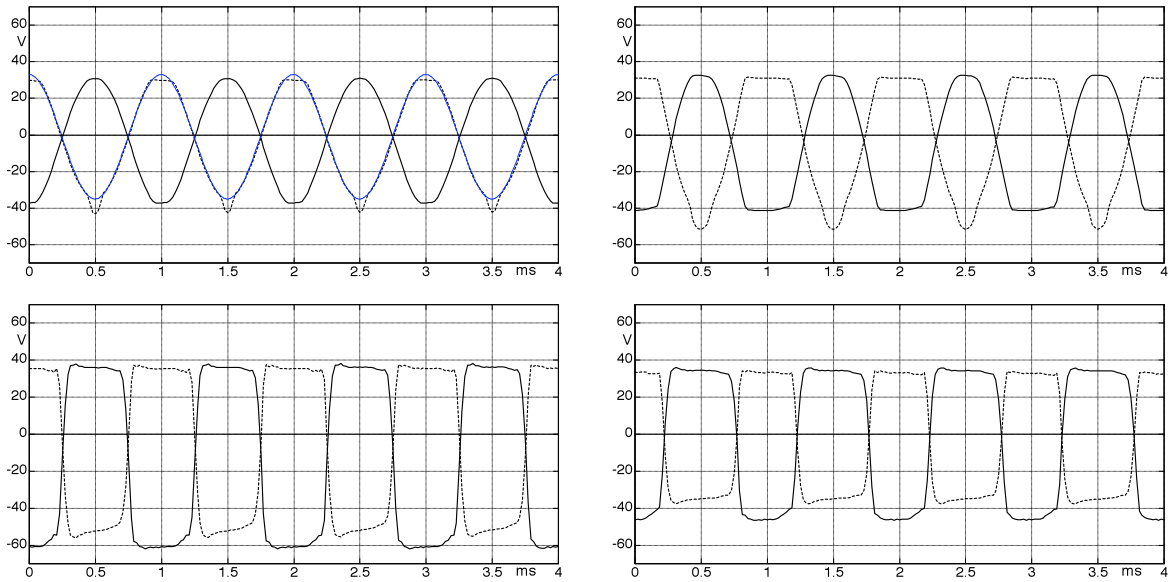


Fig. 10.4.7: Cathodyne-stage with load; AC-component. Plate-voltage (----), cathode-voltage (——).
The bottom right-hand picture shows the situation after longer-term overdrive.

10.4.3 Differential amplifier (long-tail)

This type of circuit unites two different basic tube-amplifier-concepts: the first tube works in a common-cathode configuration with current-based negative feedback; the second tube operates in common-grid configuration and is driven by the first tube via the cathode. In Fender-history, the differential amplifier represents the final step in series of developments: paraphase (1946 – 1951), paraphase with negative feedback (1951 – 1954), cathodyne (1955 – 1957), and differential amplifier (from 1956). Other manufacturers, such as e.g. VOX (1958) or Marshall (1962) that start amplifier production more than a decade later than Fender, use the differential amplifier right from the start.

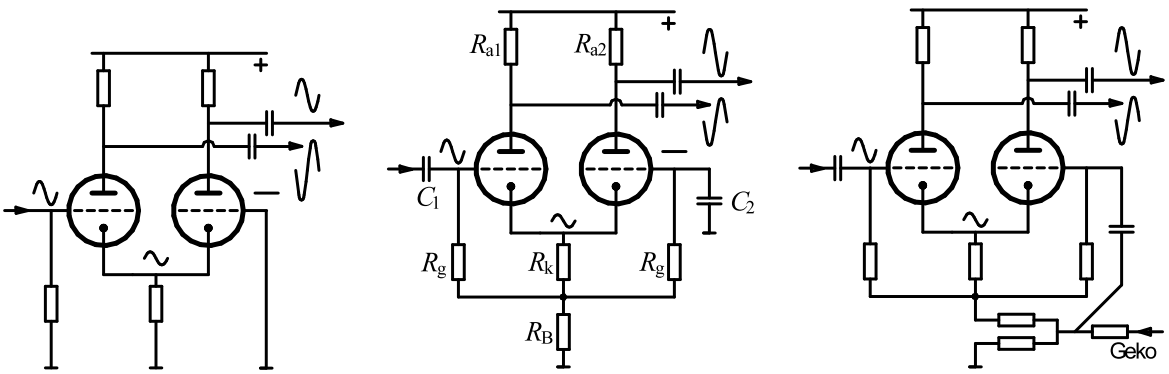


Fig. 10.4.8: Differential amplifier with negative feedback via the cathode (Geko = negative feedback).

The left section of **Fig**. **10.4.8** shows the basic arrangement of the differential amplifier. Driving the left tube with an AC-voltage changes its plate- and cathode-currents and thus creates a voltage-drop at the plate- and cathode-resistors. The cathode-voltage of the left tube changes the drive-voltage of the right-hand tube, as well, and also here causes changes in the plate- and cathode-currents (common-grid-circuit). An **example**: if the grid-voltage (defined against ground) of the left tube rises by 2 mV, the cathode-voltage increases by 1 mV. Its grid-to-cathode-voltage therefore has increased by 1 mV while the grid-to-cathode-voltage of the right tube has decreased by 1 mV. For identical transconductances of the tubes, the result would be plate-voltages of the same amplitude but opposite phase. Text-books like to use this example – but it does have a flaw: the sum of the *changes* of the plate-currents would be zero, and the cathode-potential would remain constant, i.e the right tube would not receive a drive signal. We can introduce a small correction to make the example work: the left grid-potential rises by 3 mV, the cathode-potential by 1 mV, the plate-voltages are of opposite phase … but not of the same amplitude anymore! Given typical component values, the AC-voltage-gain of the right tube would be only about half of that of the left tube, plus it would be rather strongly dependent on individual tube data. For this reason, the cathode-resistor is increased. This reduces the gain of the two tubes, but also the dependency on the individual tube (current-based negative feedback). The middle section in Fig. 10.4.8 shows such a circuit (VOX AC-30), the right-hand section also presents an input for a negative-feedback (NFB) loop that would be closed via a line from the output transformer (Marshall, Fender from 1956).

For the typical tube for the differential amplifier, Fender uses the **12AX7** (7025, ECC83) first but then changes (in the Blackface era) to the lower-impedance **12AT7** (ECC81). VOX uses the ECC83 (12AX7); Marshall does, as well.

DATA-SHEET SPECIFICATIONS: Internal impedance = 30 kΩ (ECC81) and 63 kΩ (ECC83).

An exact analysis of the differential-amplifier circuit shows that the voltage gains of the two tubes are different, despite the negative feedback. In a typical Fender configuration (Pro Amp AA763: R_a = 100 kΩ, R_g = 1 MΩ, R_k = 470 Ω, R_B = 27 kΩ), this difference is about 7%. It is likely that for this reason one of the plate-resistors (R_{a1}) was changed to 82 kΩ in a later model (Pro Reverb AA 165). For the following variant (AB 668), the plate-resistors are again equal in value but have merely 47 kΩ – and this arrangement remains for some time. VOX uses two resistors of equal value (and completely dispenses with any overall negative feedback!), while Marshall mostly employs the 82k/100k-pairing, and a frequency-dependent overall negative feedback.

The grid-resistor R_g of the first tube usually has 1 MΩ; this value was probably also seen as the input impedance. With a 10-nF-coupling-capacitor (e.g. Fender Twin 5F8A), a high-pass cutoff-frequency of 8 Hz would result – that is very low for a guitar amp but certainly compatible with the HiFi-preachings of the day. The negative feedback (R_B), however, does not only decrease the voltage gain, but it also increases the input impedance (bootstrap) from 1 MΩ to 2 MΩ, pushing the cutoff-frequency to a subsonic 4 Hz. That would more than suffice even for a bass amplifier, and indeed the 5F6-Bassman includes the 20-nF-coupling-capacitor, as well. But: a few years later the 6G6-B-Bassman receives a coupling-capacitor of a mere 500 pF! The calculation would yield a high 160 Hz as the lower cutoff-frequency, but we must not overlook that a second negative feedback loop is operating besides the feedback via the cathode. This complicates the calculation because further phase-shifting RC-circuits are in the game, and in particular the output transformer requires consideration. We had only the schematic of the 6G6-B-Bassman and no original amplifier at our disposal so no quantitative elaborations shall be included here. Just this general statement: Fender used very different capacitances (250 pF – 20 nF) for the input capacitor (C_1) of the differential amplifier; the actual high-pass cutoff-frequencies of these different circuits should be measured and not just calculated from the schematics. By the way: C_1 is 47 nF in the AC-30 and 22 nF in the Marshall.

In **Fig. 10.4.9**, the grid-voltages of a Fender Super-Reverb are shown for three different drive levels. For a small drive level, the two signals show minor differences in their amplitudes but at high drive levels there is a significant asymmetry. We could ignore the differences in the limiting towards negative voltages because the respective output tube will be in cut-off state anyway; however, due to the differences in the DC-component in the two drive-signals the two coupling-capacitors are polarized differently, leading to different duty-cycles in the plate-currents of the power amplifier. In Chapter 10.4.4, we will take an in-depth look at this asymmetry caused by the grid-current.

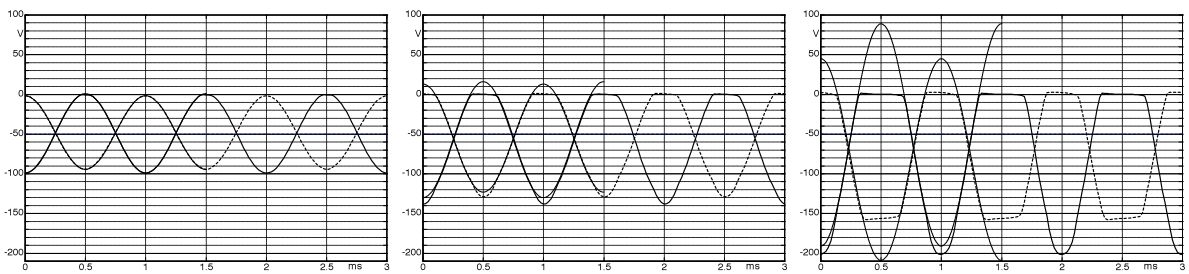


Fig. 10.4.9: Measurements at the differential amp of a Fender Super-Reverb (AB-763, negative feedback deactivated). Power-tube bias = –50V. Grid-voltage of the 1st power tube (V7 = ——), and of the 2nd power tube (V8 = ----). On the left, undistorted cosine-oscillations are shown for comparison.

10.4.4 Half-wave anti-symmetry

Each of the two power tubes generates both even-order and odd-order distortions; however, as the two separately generated half-waves are superimposed, the even-order distortions cancel each other out (half-wave anti-symmetry, Fourier-transform). This would be the ideal scenario that would require:

- the output voltages of the phase-inverter to be as similar as possible,
- the power-tubes to be as similar as possible (i.e. paired),
- the primary windings of the output transformer to be as equal as possible.

Classical amplifier technology offers solutions for signal amplification with as little distortion as possible, and regards the minimization of the even-order distortion as an advantage of the push-pull power stage. We will not investigate here whether even-order distortion (i.e. k_2, k_4, etc.) sounds good or bad in a *guitar*-amplifier – that would be a subject for psychoacoustics (Chapter 10.8). The following analyses will focus on the question how far the distortion-minimization is in fact successful.

Within the push-pull Class-B power stage (Chapter 10.5.3), the signal is spit into two parallel, opposite-phase signal paths – each power tube amplifies only one half-wave. The superposition towards the overall signal happens in the output transformer (**Fig. 10.1.10**). Ideally, no error at all would occur in this process with all spectral lines except the 1st harmonic cancelling each other out in the superposition. Of course, the splitting and re-composition will not work flawlessly in reality, and non-linear distortion will appear.

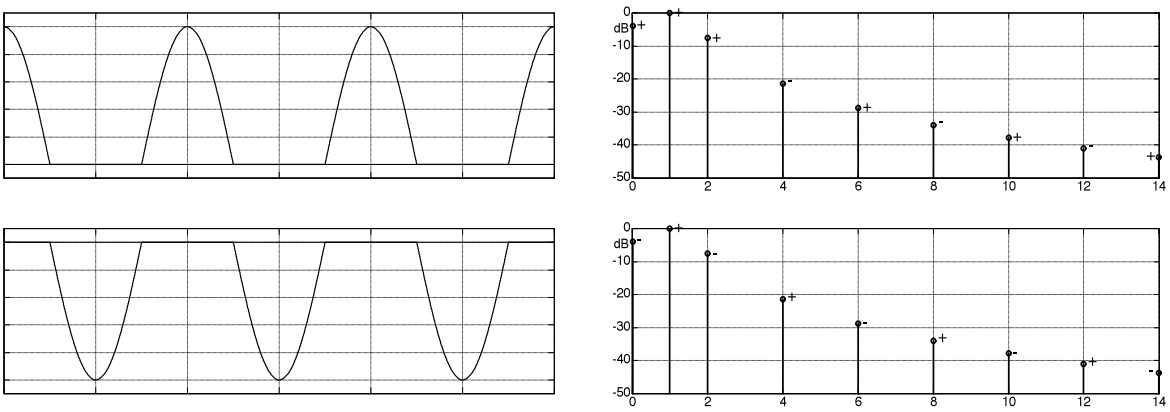


Fig. 10.4.10: Time functions (left) and spectra of the half-wave signals. The signs of the Fourier-components are the same only for the 1st harmonic, and consequently only this component remains after the addition.

An obvious error results from the unequal amplification of the two half-waves (**Fig. 10.4.11**). The compensation of the even-order harmonics is incomplete and even-order distortion remains ($k_2 \approx 8\%$ in the picture).

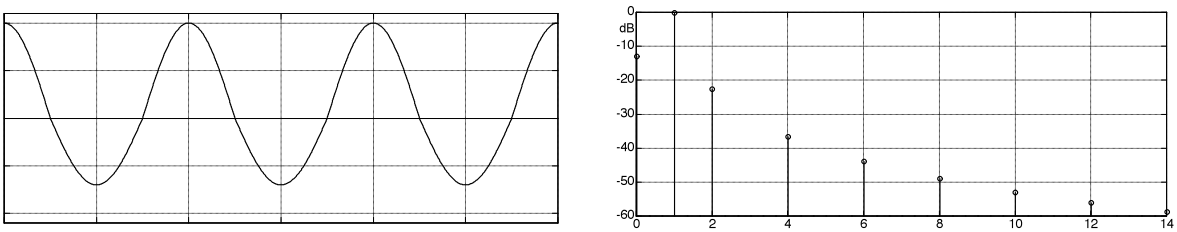


Fig. 10.4.11: Time function and spectrum of a signal with different amplification of the two half-waves.

For the time function shown in Fig. 10.1.11, the two half-waves have different amplitudes – they are, however, not half-wave anti-symmetric. **Half-wave anti-symmetry** stands for a time-periodic signal repeating itself, with inverted sign, after half a signal-period: $u(t) = -u(t + T/2)$. From the rules of the Fourier-transform, it directly follows that such a signal can only contain odd harmonics. Consequently, only distortion products of odd order (k_3, k_5, k_7 etc.) can be generated as long as the transmission characteristics of the two half-wave transmission branches are equal. "Asymmetry♣", however, already starts in the **phase-splitter stage** for the drive signals. The two gains in the paraphase-branches (Chapter 10.4.1) are as different as the two tube-systems in the double-triode – that's why quite early on the doctor (or rather Leo F.) ordered a negative-feedback loop. Cathodyne-circuit and differential amplifier show much less dependency on the individual tube data, and in fact they *could* deliver two signals equal in amplitude and opposed in phase with sufficient precision – but only as long as there are negligible grid-currents. Why do we find asymmetries already in the schematics, why do the gain factors differ for the two half-waves, even for ideal tubes? Answers have been and remain speculative:

1. the designers of early circuits were not yet that well versed in electronics, and later the archetypes continued to be simply (and indiscriminately) copied.
2. these intentional "asymmetries" were supposed to give a special sound.
3. these asymmetries were supposed to correct other asymmetries in the circuit.
4. guitar amplifiers are no instrumentation devices; high accuracy was not that important.

Ad 1: This assumption cannot entirely be brushed off. Leo Fender's explanations regarding magnetism are … well, to be fair … they're what you would expect given that he was originally trained as a bookkeeper (one with aspects of a genius, without a doubt). But early on improvements creep into the circuits (whoever developed them): the paraphase circuit with negative feedback appears around 1954 in the Fender Deluxe i.e. it was desirable that the asymmetries created by the tube-variances didn't take over too much. Balancing a power amplifier can be done without any grand network-analysis: with an oscilloscope and a resistor-decade you come already pretty far, and such equipment was probably available even in the labs (or workshops, rather) of the early protagonists.

Ad 2: That is an alluring thought but it asks for a bit of dispute. On the one hand: your regular musician (or customer) will not be able (or willing) to un- and re-solder resistors after each tube-change. If the asymmetry mentioned above were decisive for the sound, it would be purely accidental because no circuit will totally equalize out the tube variances (in particular those of the power tubes). We would have a contradiction to the objective of achieving a *special, sought after* sound. On the other hand: this is exactly why musicians will choose that one best-sounding amp from a group of 5 Deluxes (or Super-Reverbs, or Twins …). Understandably, you are not allowed to ask whether this amp can be switched on ever again at all (so that the tubes may not age, and to preserve the incomparable sound). "Just buy some more NOS-tubes" – that's what advertising will recommend.

Ad 3: there may be some truth to that, was well – possibly connected to 1. A designer discovers that the phase-splitter stage needs to work in an un-balanced mode to obtain a fully symmetric signal at the speaker output. Maybe the output transformer has a special asymmetry? Not because the winding-machine has failed to count correctly, but because there are slightly different (magnetic) coupling factors. Indeed, that may be compensated via the phase-splitter stage – but of course only as long as the transformer data always remain the same.

♣ we could call this "un-anti-symmetry" just as well

Ad 4: Of course, every designer gets to the point where additional effort is not sensibly warranted anymore in view of the costs additionally incurred. Although: a 100-kΩ-resistor costs just as much as an 82-kΩ-resistor. Following-up the development of resistor-values in the phase-splitter over the years, we easily recognize the fight for the "optimum solution" (Chapter 10.4.3). Overall-negative-feedback approaches that include even asymmetries in the magnetic fields bear testimony to the desire for reducing non-linearity as much as at all possible. There are counterexamples, though, such as the AC-30 with a power amp that must make do without any negative feedback – and this surely not just because of the cost-factor.

So, there we are. As already mentioned; the answers were always and remain speculative. Maybe the following mixture was a typical situation: the expressed objective was a symmetry as good as possible, ergo little k_2, and so the prototype in the workshop was modified until the result was something the designer could be proud of – and hopefully sounded good, as well. And off to production … the next project awaits. Creating statistics about parameter variances was likely to be as popular in the 1950's as it is today – and it was apparently not necessary, either.

Unless we are checking out a completely out-of-control paraphase circuit, the tolerances ("un-anti-symmetries") occurring in a typical phase-splitter stage for **small-signal operation** are rather insignificant, especially compared to the idiosyncrasies in the **large-signal behavior**. In order to get from the high plate- (or cathode-) potential to the low grid-potential of the two power tubes, every usual phase-splitter stage uses two coupling capacitors (**coupling-C**'s) carrying the two signals driving the power-tubes. The coupling-C *"separates the DC-component"* and carries a constant DC-voltage across it – tells us theory, anyway. It ain't so! As distortion (not actually forbidden in guitar amps!) occurs in the output tubes, the latter experience a non-negligible grid-current which changes the DC-voltage across the coupling-C's and thus also the operating point of the output tubes.

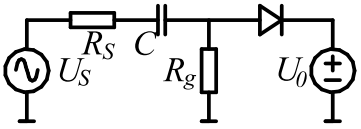


Fig. 10.4.12: Simple model-circuit to simulate grid-currents.

Fig. 10.4.12 presents a simple circuit enabling us to discuss the basic behavior in case of occurrence of a grid-current. U_S is the signal-source (i.e. the tube of the phase-splitter) with its internal impedance R_S, C is the coupling capacitor. R_g stands for the gird-resistor of the output tube (e.g. 220 kΩ); the non-linear input impedance of the output tube is modeled by the diode and the DC-voltage source (e.g. U_0 = 20 V). As a first step, it is conducive to assume the AC-voltage source not to have an additional DC-offset.

As long as the amplitude of the AC-voltage U_S is smaller than U_0, the diode (thought to be ideal) is in blocking mode. Only a minimum AC-voltage and no DC-voltage is found across the coupling-C (assuming operation significantly above the high-pass cutoff-frequency). However, as the AC-amplitude $\hat{U}_S$ rises above the DC-voltage U_0, the diode starts to conduct and limits the signal across R_g. The diode now carries an impulse-shaped current flowing only in one direction and thus having a mean value different from zero. We could also say: a DC-free AC-current with superimposed DC-current flows through the diode. The DC-current-part can, however, not pass through the capacitor and has to flow in total through R_g, generating a (negative) voltage across the resistor. The source (U_S) remains free of any DC-voltage (stiff voltage source), but across R_g we get a DC-voltage, and consequently the DC-current polarizes the coupling capacitor.

This **polarization** of the coupling capacitor is a non-linear process that could be described via a non-linear differential equation. As a simplification, we can also look at the final process-state and assume the polarizing voltage across the coupling-C to be constant (but dependent on the drive level). **Fig. 10.4.13** shows several corresponding time-functions: the amplitude of the source voltage is 35 V in both sections of the figure; in the left-hand section the signal is only limited, and in the right-hand section it is additionally shifted towards negative values. This voltage-shift is the polarization-voltage across the capacitor.

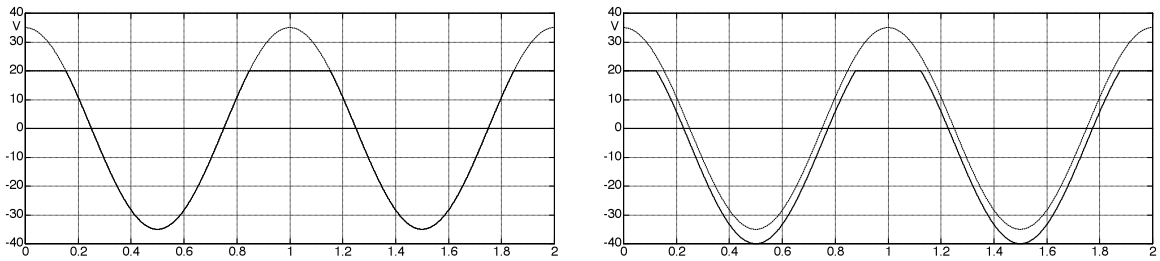


Fig. 10.4.13: Potential-shift due to grid-current in the output tubes. Left: AC-voltage limited to merely 20 V; right: AC-voltage limited and shifted (capacitor-polarization).

Only for strong drive levels, or for overdrive, any relevant grid-current starts to flow in the output amplifier, and only these currents lead to a re-charging of the coupling capacitors, and thus to a shift in the operating points of the output tubes. In **Fig. 10.4.14**, we see this polarization voltage given for two different series-resistors as a function of the signal amplitude.

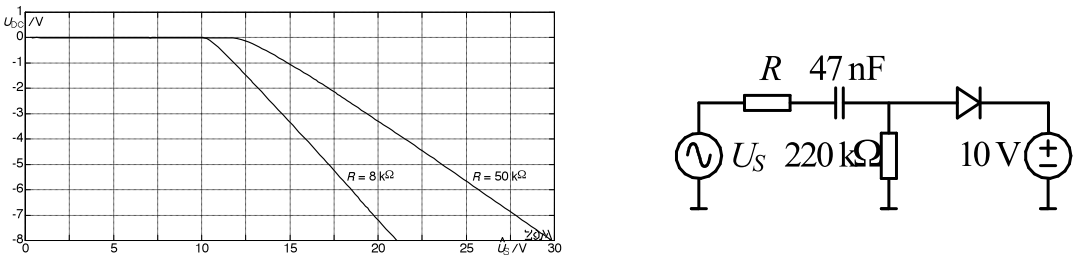


Fig. 10.4.14: Average grid-voltage-bias U_{DC} in dependence on the drive-voltage-amplitude (model).

In contrast to this model, we find – in the real-world push-pull power amplifier – a voltage across the capacitors even without any drive signal. This is the difference between the plate-voltage (e.g. 250 V) and the grid-bias voltage of the output tube (e.g. -50 V). In **Fig. 10.4.15** the mean value of the grid voltage of the output tubes is shown as a function of the drive level. As mentioned above, the grid becomes more negative as the grid-current increases. For the 2nd output tube (V 8), there are potential shifts already at small drive levels. This is not due to any grid current, but caused by shifts in the operating point of the differential amplifier.

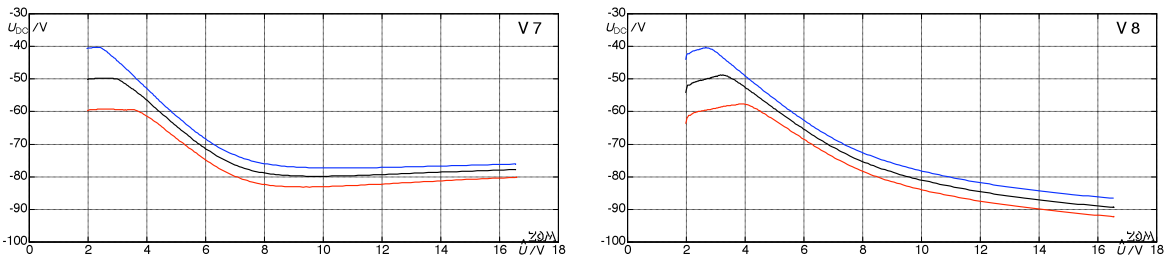


Fig. 10.4.15: Fender Super-Reverb, grid-bias-voltage of output tubes (mean); 3 different operating points. Drive voltage (abscissa) is the grid-voltage of the left-hand differential-amplifier tube.

The mean values of the plate-voltages of the phase-splitter do not remain constant as a drive-signal is applied; they shift even for moderate levels (**Fig.10.4.16**). Consequently, the polarization-voltage levels of all four capacitors change – with very different time-constants taking effect. For example, $C_2 = 0.1\ \mu F$ is recharged via $R_g = 1\ M\Omega$, resulting in $\tau = 0.1$ s. The capacitors branching off the plates need to be re-charged, as well, and thus re-charging currents flow through the grid-resistors (not shown in the figure) of the output tubes. Consequently, the operating points of the output tubes are shifted due to two mechanisms: the potential shifts in the differential amplifier, and the grid-currents flowing in the output tubes.

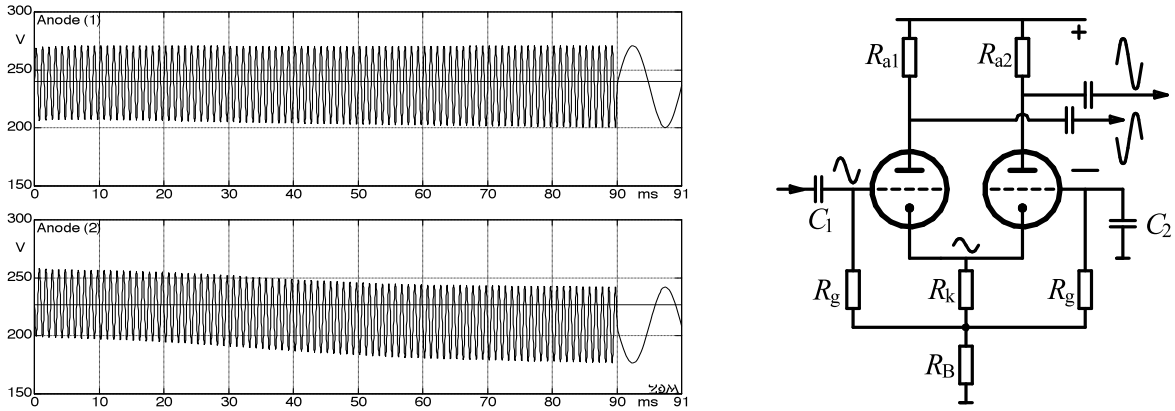


Fig. 10.4.16: Shift of the operating point in the differential amplifier of a Super-Reverb (negative feedback deactivated). The mean-value of the plate-voltage for the right-hand triode shifts towards lower voltages.

We can see from **Fig. 10.4.17**, that these drive-dependent re-charging processes in the differential amplifier do not happen in a symmetrical fashion: for small drive-levels, both mean values of the plate-voltages decrease, while for strong drive-levels the mean plate-voltage of tube 1 increases while the plate-voltage for tube 2 decreases. Switching off the drive signal makes the grid-voltage at the 1st output-tube (V7) jump to more negative values while this jump is to more positive values for the other output tube (V8). Consequently, there will be a superposition of interferences of very low frequencies on top of the useful signal. We could ignore this because neither the output transformer nor the loudspeaker nor the hearing system is susceptible to such low-frequency excitation – still, we must not generally neglect these side-effects because corresponding operating-point shifts can lead to envelope modulation and time-variant non-linear distortion.

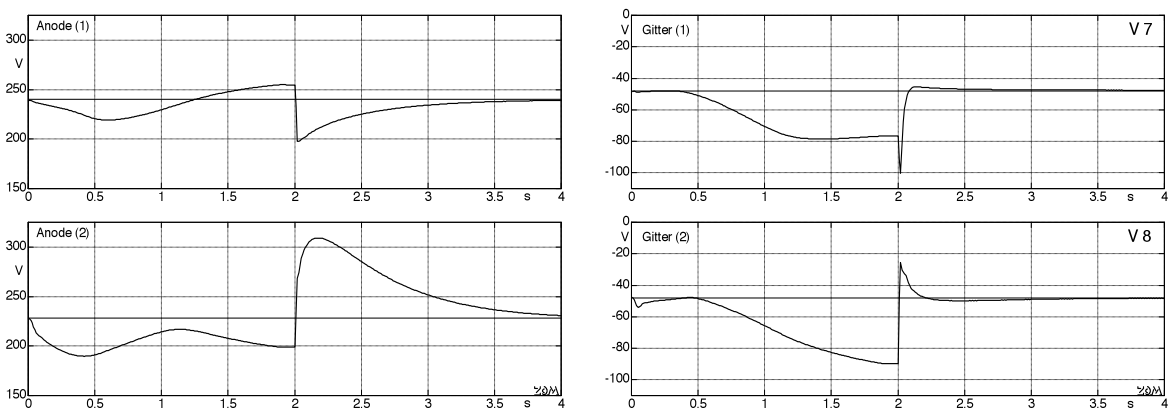


Fig. 10.4.17: Mean values of the voltages at the plates of the differential amplifier (left) and at the output tube grids. During $0 < t < 2$ s, the signal level rises by 20 dB, at $t = 2$ s the signal is shut off. Super-Reverb.

Fig. 10.4.18 shows corresponding loudspeaker voltages of a Super Reverb that had its overall negative-feedback loop (via the output transducer) deactivated. **In the left-hand part of the figure**, a 1-kHz-tone that overdrives the power-amplifier is switched on at $t = 0$. At $t = 100$ ms, the level of the tone is reduced♣ by 20 dB which makes the loudspeaker voltage collapse for a short time. We should not dramatize such effects (compare to the post-masking effects in the hearing system) but we should not generally ignore them, either, because there may be individual cases with longer time constants, and because music does not really consist of exclusively 20-dB-jumps. **In the right-hand section of the picture**, the loudspeaker voltage is depicted for almost full drive and for overdrive. Caused by the potential shifts connected to the grid-current, saddle-point-shaped distortions appear for overdrive-operation at the **zero-crossings**. These distortions cannot be traced to insufficient biasing or output-transformer saturation, as it is sometimes surmised in literature.

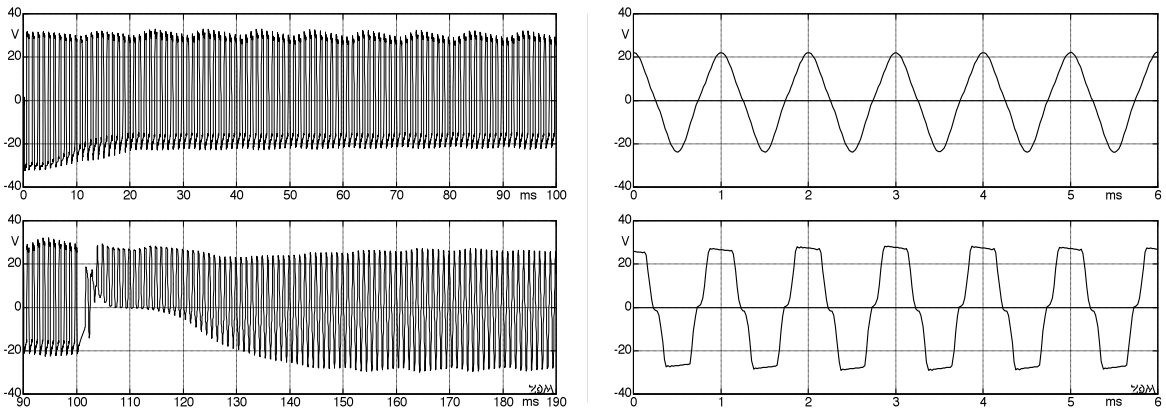


Fig. 10.4.18: Super-Reverb, loudspeaker-voltage (overall feedback-loop deactivated).

The saddle-points (also termed crossover-distortion) appearing at the zero-crossings occur if the half-waves, separately processed by the output tubes, cannot be joined precisely enough. The superposition does not work sufficiently with the tube-characteristics moving apart due to the shifts of the mean voltage-values (**Fig. 10.4.19**). For supplements, see Chapter 10.5.8.

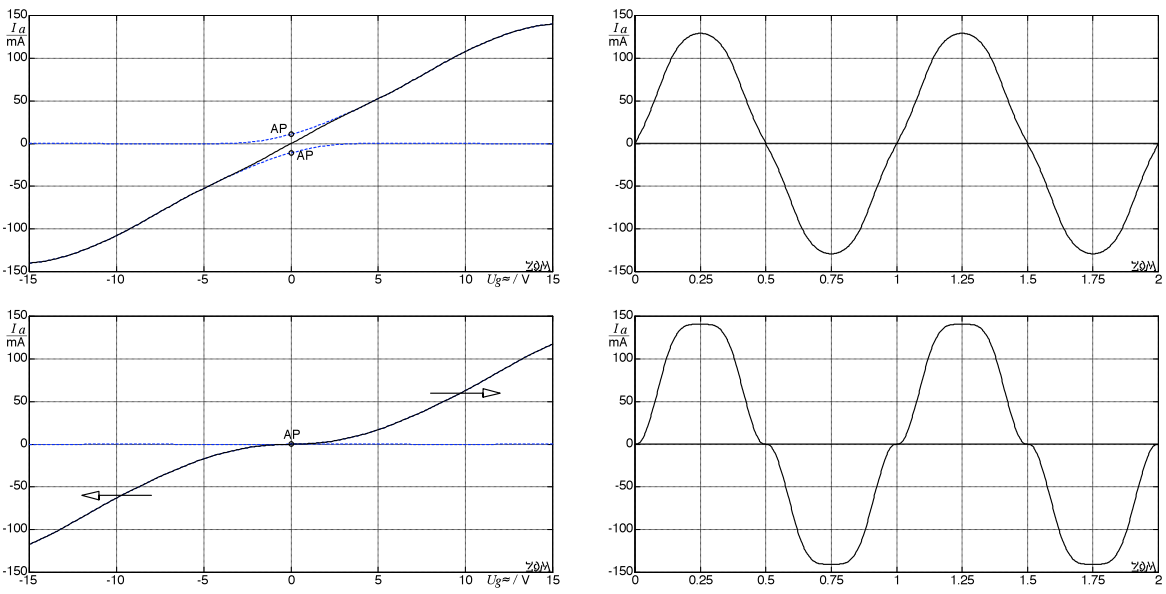


Fig. 10.4.19: Dynamic (drive-level dependent) crossover distortion (compare to Chapter 10.5.8).

♣ The power-amplifier still remains overdriven

