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.

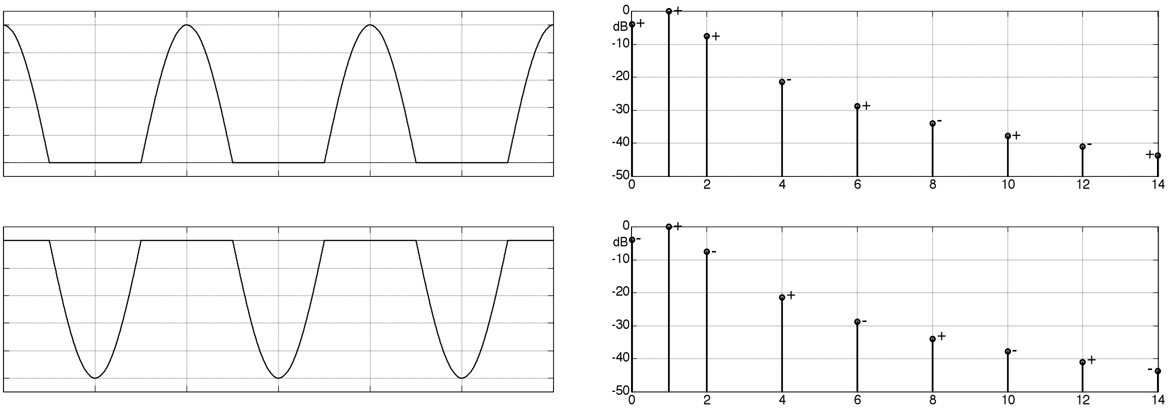

Fig. 10.4.10: Time functions (left) and spectra of the half-wave signals. The signs of the Fourier-components are the same only for the 1st harmonic, and consequently only this component remains after the addition.

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).

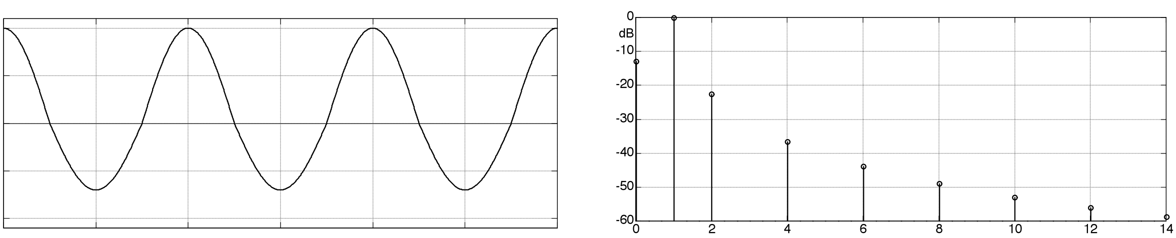

Fig. 10.4.11: Time function and spectrum of a signal with different amplification of the two half-waves.

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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.

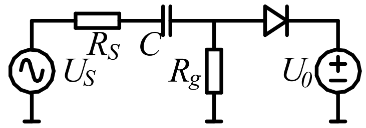

Fig. 10.4.12: Simple model-circuit to simulate grid-currents.

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.

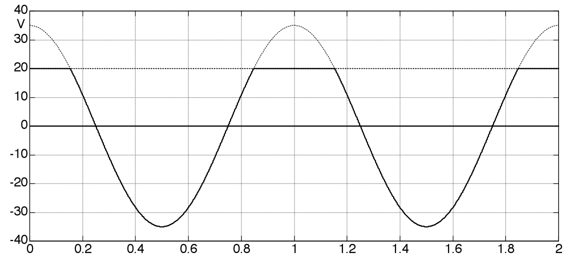

Fig. 10.4.13: Potential-shift due to grid-current in the output tubes. Left: AC-voltage limited to merely 20 V; right: AC-voltage limited and shifted (capacitor-polarization).

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.

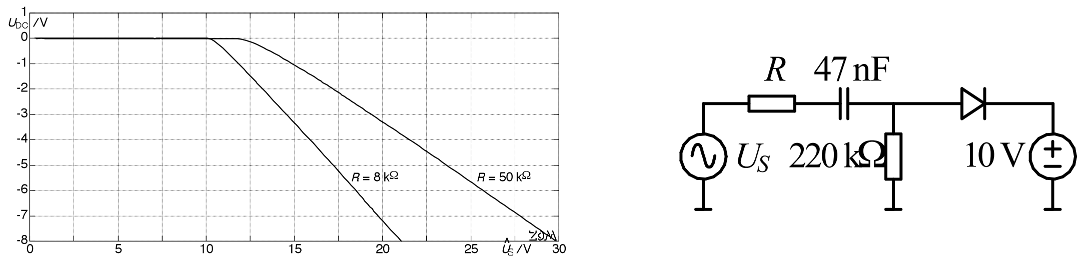

Fig. 10.4.14: Average grid-voltage-bias U_{DC} in dependence on the drive-voltage-amplitude (model).

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.

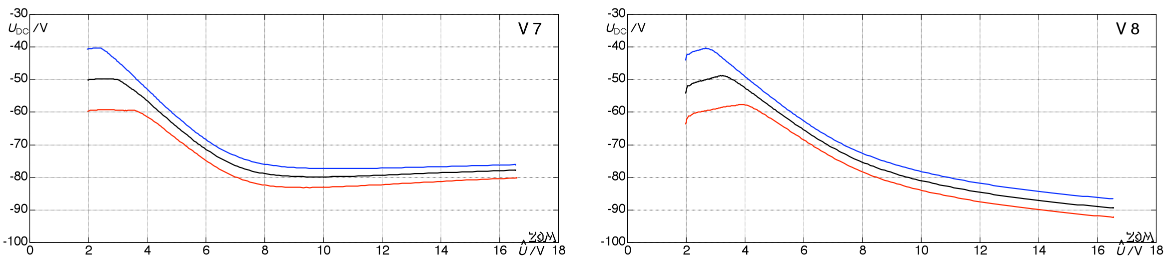

Fig. 10.4.15: Fender Super-Reverb, grid-bias-voltage of output tubes (mean); 3 different operating points. Drive voltage (abscissa) is the grid-voltage of the left-hand differential-amplifier tube.

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 μF is recharged via R_g = 1 MΩ, resulting in τ = 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.

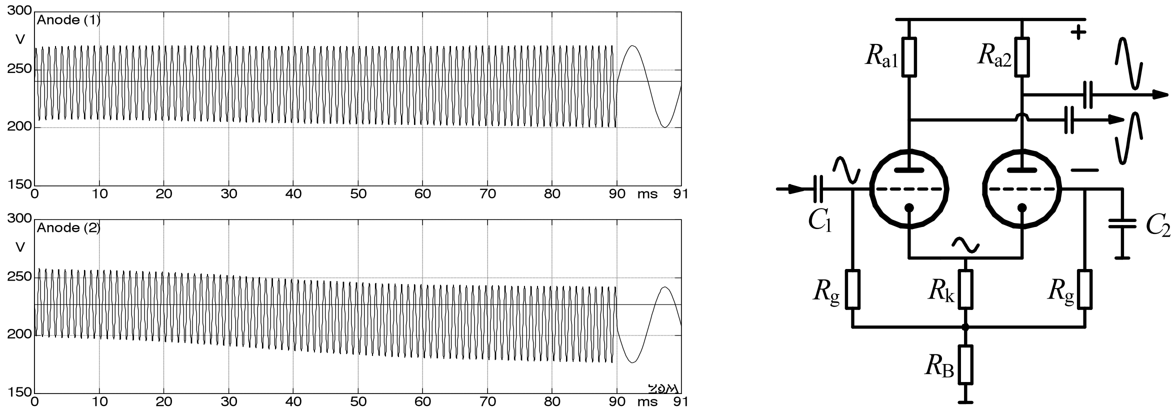

Fig. 10.4.16: Shift of the operating point in the differential amplifier of a Super-Reverb (negative feedback deactivated). The mean-value of the plate-voltage for the right-hand triode shifts towards lower voltages.

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.

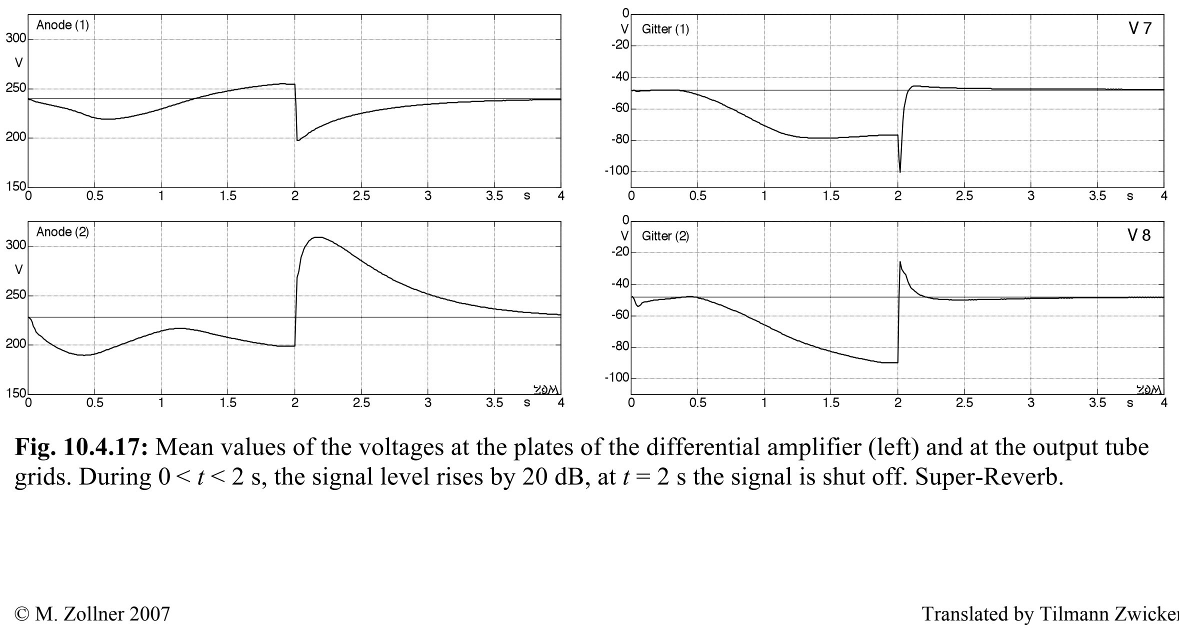

Fig. 10.4.17: Mean values of the voltages at the plates of the differential amplifier (left) and at the output tube grids. During $0 < t < 2$ s, the signal level rises by 20 dB, at $t = 2$ s the signal is shut off. Super-Reverb.

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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.

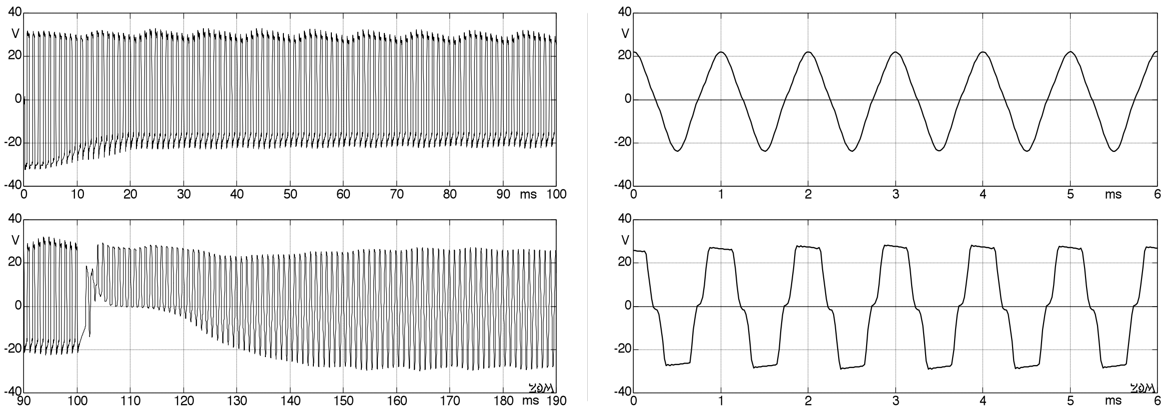

Fig. 10.4.18: Super-Reverb, loudspeaker-voltage (overall feedback-loop deactivated).

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.

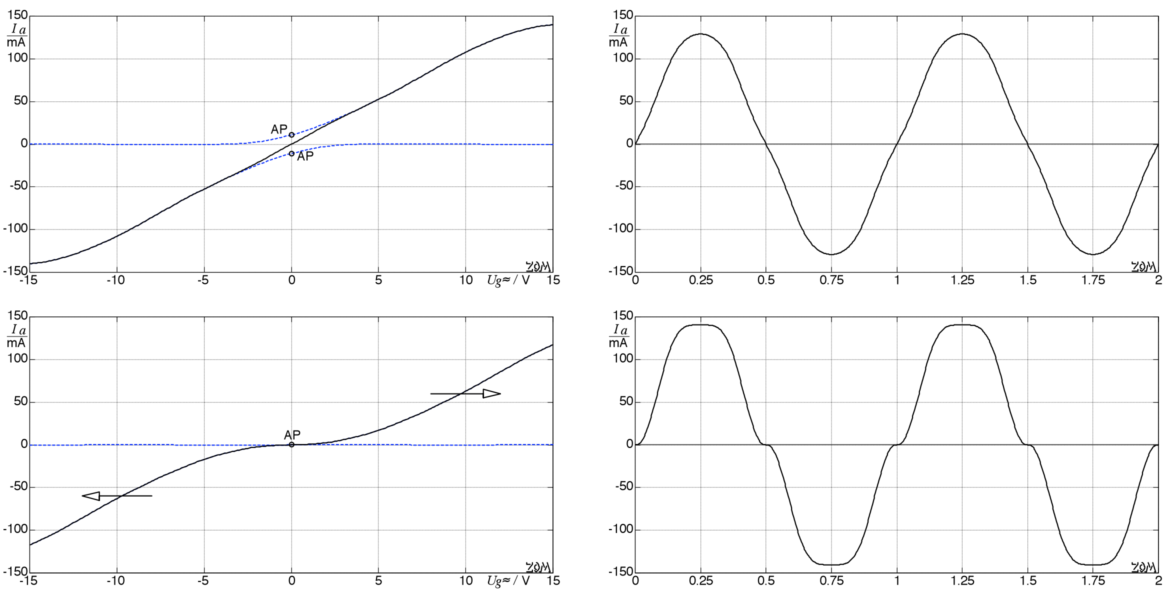

Fig. 10.4.19: Dynamic (drive-level dependent) crossover distortion (compare to Chapter 10.5.8).

♣ The power-amplifier still remains overdriven