10 Guitar amplifiers

Electric guitars per se radiate only very little sound – to be decently heard, they require special amplifiers and loudspeakers. Indeed, one is well advised to better regard amplifier and loudspeaker as an integral part of the musical instrument: in a manner of speaking, the electric guitar does extend up to the loudspeaker. Guitar amps create distortion and usually feature a frequency dependent transmission-factor; the attached speakers create distortion, as well, and do show an uneven frequency-response – and only if, on top of everything, the loudspeaker enclosure has pronounced resonances will the guitar player "be satisfied". However, there are also ugly, buzzy distortions, and not every resonance or kink in the frequency-response will sound good. We have seen innumerable attempts to improve the primitive circuits of the first guitar amplifiers – alas, in many cases the circuits may have improved but the sound got worse. Textbooks on circuit-design teach about avoiding the non-ideal characteristics of circuits; for example: how negative feedback will reduce the nonlinear distortion of the power-amplifier. Some famous guitar amps, however, achieve their great sound especially because indeed they dispense with all negative feedback in the power section – the VOX AC-30 being a most prominent example. On the other hand, to conclude that to this day science has failed to understand the functionalities of a tube amplifier – that would be far from the truth. Indeed, systems-theory, circuit design and instrumentation technology are powerful and successful areas in electronics … the issue here is the definition of the task at hand. It must precisely not be the aim to "linearize" the frequency-response (i.e. to render the transmission-factor frequency-independent), but we need to e.g. follow up the question how dents in the frequency-response influences the *sound*. It is this subjectively perceived sound that is important, not so much the physically measured sound. The question whether 2nd-order distortion makes an electric guitar sound better than 3rd-order distortion is not included in the tube manual. Whether such distortion should happen in the pre-amplifier or rather in the power stage is not discussed, either. It is in particular this interaction of the individual system parts that renders the circuit-analysis and -design so complex and difficult. The simplifying description as LTI-system can be only a first step and needs to be followed up by further steps. Thus, circuit analysis of guitar amps requires significant effort; the following elaborations therefore confine themselves to a few generic circuits[1].

10.1 Preamplifier

In circuit design, the circuit section grouped around an amplifier tube is i.a. designated an amplifier *stage*. Typical are: preamplifier, tone-filter, phase-inverter, power amplifier, and power supply. Each of these partial circuits contributes to the sound of an amplifier, or, rather to its transfer characteristic. In the preamplifier (or input amplifier), the preamp tube amplifies the signal by a factor of 20 – 50. During the first two decades of amp-history, large octal tubes (tubes with an 8-pin socket) were used, but in the mid-1950's the smaller 9-pin noval tubes were introduced to the booming amp market. Especially the high-gain 12AX7 (ECC83, 7025) has established itself as a standard still recognized over half a century later.

[1] Translator's note: In this chapter, measurements taken on a number of typical amplifiers are shown. These amplifier specimen were for the most part newly built exactly according to the schematics and layout of the historic originals, i.e. they were in a state as if they had just come off the assembly line (comparable to „N.O.S."). In one case (VOX AC-30) an original form the 1960's was available and used; this amp had been restored to its original working condition using new components of accurate values where needed. **N.B.: the Fender Super Reverb used for the measurements had an output transformer with both 2-Ω- and 8-Ω-outputs – this made comparisons with the other amplifiers easier.**

10.1.1 Preamplifier tube

Fig. 10.1.1 juxtaposes the circuit of a triode as it is typically used in input stages of guitar amplifiers, and an N-channel-JFET circuit. These circuits are not equivalent but the comparison will assist the solid-state-expert to easily access to the world of tubes. The three electrodes of the **triode** are designated **cathode**, **grid**, and anode or, more commonly used, **plate**. They correspond to source, gate, and drain in the JFET. Unlike with the FET, the tube requires a heater current (about 0,3 A at 6,3 V) that normally is not shown in the schematic. Tubes operate using very high supply-voltages (U_B = 200 – 400V) i.e. 10 times the value for the FET. On the other hand, the currents flowing in the plate- and in the drain-circuit are comparable: for input stages, they amount to about 1 – 2 mA. The voltage between grid and cathode (gate and source) constitutes the control quantity; for small drive levels, the input (grid, gate) is of very high impedance – i.e. the grid-(gate)-current through the grid-resistor (gate-resistor) R_g, is negligible. The cathode-(source)-current therefore is equal to the plate-(drain)-current. For a cathode-(source)-current of 0,8 mA we find a voltage of 1,2 V across the cathode-resistor R_k (source-resistor R_s), and consequently the control voltage U_{gk} (U_{gs}) amounts to –1,2V as long as the input voltage U_e remains at zero. For guitar amps, the input impedance R_e is often 1 MΩ, and the series resistor R_g often amounts to 34 kΩ (two 68-kΩ-resistors in parallel), while the plate-resistor R_a will be between 100 kΩ and 200 kΩ.

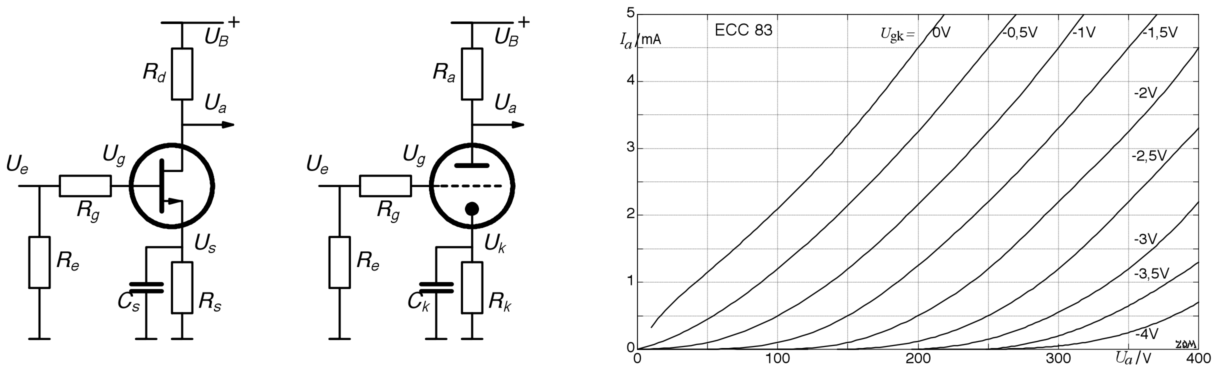

Fig. 10.1.1: Input-circuit of a tube amplifier (center) compared to a FET-amplifier (left). The right-hand picture shows typical characteristics (data-sheet of the double-triode-tube ECC83). The term “control voltage” is used in various ways – here, the grid/cathode-voltage and the gate/source-voltage is meant.

At the operating point (i.e. without drive signal, $U_e = U_g = 0$), the control voltage U_{gk} (U_{gs}) is (at e.g. -1,2 V) negative for both circuits. Positive (i.e. less negative, e.g. -1V) control voltages make both amplifying elements conduct better: plate-(drain)- and cathode-(source)-currents rise. Without any cathode-(source)-capacitor, the cathode-(source)-voltage would consequently increase and thus counteract the drive signal, effectively causing negative feedback and decreasing the gain. Since early guitar amps had to make do with few tubes, high gain was required and such a negative feedback was uncalled for. Therefore, the cathode-resistor was bridged via an electrolytic **capacitor** (typically 25 µF) eliminating any AC-voltage at the cathode (within the relevant frequency-range): C_k acts as an AC-short. With increasing input voltage U_e, the plate-current rises, and this increases the voltage across the load resistor R_a such that – for a constant supply-voltage U_B – the output voltage U_a decreases. An AC-voltage at the input will cause an amplified, opposite-phase AC-output-voltage shifted by a constant DC-voltage (e.g. 250V – 100kΩ·0,8mA = 170V). The achievable voltage gain depends significantly on the type of tube: the often-used **ECC83** allows for an **AC-voltage-gain of about -50**.

The AC-voltage at the plate is out-of-phase with the grid-AC-voltage – which is why occasionally we find a **“minus”-sign** (v = –50) in the gain specification. It would also be possible to define the voltage gain as the quotient of two RMS-values: now we would always get a positive gain-factor, e.g. v = +50 (RMS-values are always positive). Still, even for positive gain specification, the plate- and the grid-ac-voltages will remain out-of-phase – at least for the common-cathode-configuration (cathode ac-connected to ground) that is ubiquitous in pre-amplification stages in guitar amplifiers.

The **voltage gain** actually obtainable with a tube circuit depends on the circuitry, on the power-supply, and on the individual tube. The **open-loop gain** (designated μ or u) given in data books characterizes a very specific operational state (no load at all at the output) that does not occur for a typical preamp stage. Both the plate resistor (also called load resistor) connected between plate and supply-voltage, and the input impedance of the subsequent amplifier stage, reduce the theoretical voltage gain to the level of the real *closed-loop gain* (often simply called *gain*). For the ECC83 (a typical preamplifier tube), an open-loop gain of μ = 100 (or –100) is given; the closed-loop gain that actually obtainable is smaller: typically, 20 … 50 may be expected. Since tube-data change their characteristics as they age, the gain does not remain constant over the years. Tube production is sometimes subject to considerable tolerances (electrode material, wiring, cathode coating, etc.), the gain of two off-the-shelf ECC83 may easily differ by 10% to 20%, and even larger tolerances are not unheard of.

To analytically describe the function of an amplifier tube, simplifications are required. Typical **models for tubes** are based on the following idealized modeling laws:

Driving a tube does not require any power (the input impedance is practically infinite); the tube is a linear and time-invariant system; the upper cutoff frequency is so high that the (low-pass-limited) signal from the guitar does not receive any additional filtering; “the” tube data are found in the tables of the data books.

None of these assumptions are, however, applicable to typical guitar amplifiers – at best, they are merely useful in the framework of a rough orientation. The following chapters are a short description of those tube-characteristics that are of particular importance for guitar amplifiers. Included are typical concepts for circuits, as well. Standard text-books [e.g. Barkhausen, Schröder, RCA-handbook] give supplemental basic knowledge. It is however vital to consider that, while the classic standard works discuss in detail and to some extent very theoretically the operational behavior of the tube, they do not mention with a single word the “abuse” (i.e. the umpteen-fold overdrive) that is regular practice in guitar amps. Modern text-books concentrate on semi-conductors and special tubes (technical tubes), and are not helpful in the context of guitar amps. Those books that in fact do discuss the idiosyncrasies of a tube-powered guitar amp are often kept rather general; they hardly offer any measurement results and rarely any theoretical calculations. In the worst case, mere assumptions are circulated as they are now found almost deluge-like on the Internet. “*The cathodyne circuit sounds much tighter than the SEPP or the long-tail because already Leo Fender introduced it in the 5E6a*” N.B.: here, it appears that this circuit sounds tighter (whatever that means) not because of any technical characteristic but because it is spiritually connected to Leo Fender …

10.1.2 Tube input-impedance

Together with the cable capacitance, the input impedance of a guitar amplifier is connected in parallel with the source impedance of the guitar pickup. Looked at in a simplified manner, the amplifier input may be represented by a high-value resistor: for guitar amps about 1 MΩ is customary. The resulting damping effect on the pickup-signal is small. If, however, the input impedance of the amp is significantly lower, an audible damping effect does happen that makes itself felt (or rather heard) as a loss in brilliance. Entirely different scenarios may occur with effects boxes (e.g. treble booster, distortion device, or wah-wah) connected between guitar and amp. Their input impedance often is rather low but this needs to be seen as part of the effect.

Aside from the regular standard input (designated "1" or "Hi"), many classic tube amps offer a second input of lower sensitivity ("2" or "Low"). Due to the smaller input impedance (typically 136 kΩ), this second input makes the guitar sound less brilliant. Also, a 50%-signal-attenuation involving a voltage divider with two 68-kΩ-resistors is included, reducing down preamplifier distortion. When the **standard input** ("1") is used, the two 68-kΩ-resistors are connected in parallel with each other, and in series with the tube input. They have the effect of a low-pass filter that however only cuts out high-frequency radio transmissions – in the audible range, the low-pass effect is insignificant.

The **input capacitance** of customary guitar amplifiers is small but not always negligible compared to the cable capacitance. For a tube amplifier, the input capacitance of the preamp-triode will be around 80 – 150 pF due to the **Miller-effect**. Depending on the wiring within the amp, further line capacitances of about 50 pF may need to be added. With the standard input-circuitry for tube amps, the guitar is galvanically coupled to the grid of the first tube – there is no coupling capacitor. Only few amps (in particular very old ones) generate the grid bias via the leakage current of the grid, and therefore separate guitar and tube via a coupling-capacitor of 10 – 20 nF. The effect of this capacitor is negligible in the framework of the linear model – the operating point of the tube in this configuration is, however, not very stable at all.

The tube grid is connected neither to the plate nor to the cathode, and since the glass container insulates very well, we could indeed surmise a tube input of very high impedance. However, while plate and cathode are not connected, there is still an electric current flowing between them. This is due to the glowing cathode emitting electrons that fly – through the vacuum in the glass container – to the positively charged plate. A flow of electron is an electrical current: negative charges flowing from the cathode to the plate make (applying the technical current direction) for a positive current from plate to cathode. The electrons travelling from the cathode land on the plate and not on the grid because the plate is charged positively relative to the cathode, and the grid negatively – for the customary **operating point**, anyway. A cathode-current of e.g. 0,8 mA (Fig. 10.1.1) flows in the absence of an input signal, and with this the grid-potential is 1,2 V more negative than the cathode-voltage. However, in the case that the grid becomes positive relative to the cathode, the electrons find two attractive landing sites: the highly positive plate and the weakly positive grid. Since the plate-surface is much larger than the grid-surface, and since the plate-voltage is much higher than the grid-voltage, most of the electrons will fly to the plate. However, a small part of them does land on the grid and causes a **grid-current**. This grid-current exits the grid as a negative electron flow, i.e. it enters the grid as technical current.

There are several reasons for the **flow of this grid-current**: finite insulation resistances grid/plate and grid/cathode, ionization of the remaining gas in the glass container (deficient vacuum), thermal grid-emission due to high grid-temperature, and the already mentioned pickup of a part of the electron cloud emitted by the cathode. The individual effects superimpose (in part with inverse signs) and result in a non-linear input characteristic; the grid-current depends on the grid/cathode-voltage U_{gk} in a non-linear fashion. For input voltages♣ U_e of above about +0,7 V (U_{gk} > –0,5 V) there will be an observable grid-current leading to a voltage across the grid-resistor R_g. Consequently, the grid-voltage U_g decreases. This effect makes itself felt especially for strongly positive input voltages: for example, we may find only about +1,2 V instead of +4 V at the grid (**Fig. 10.1.2**).

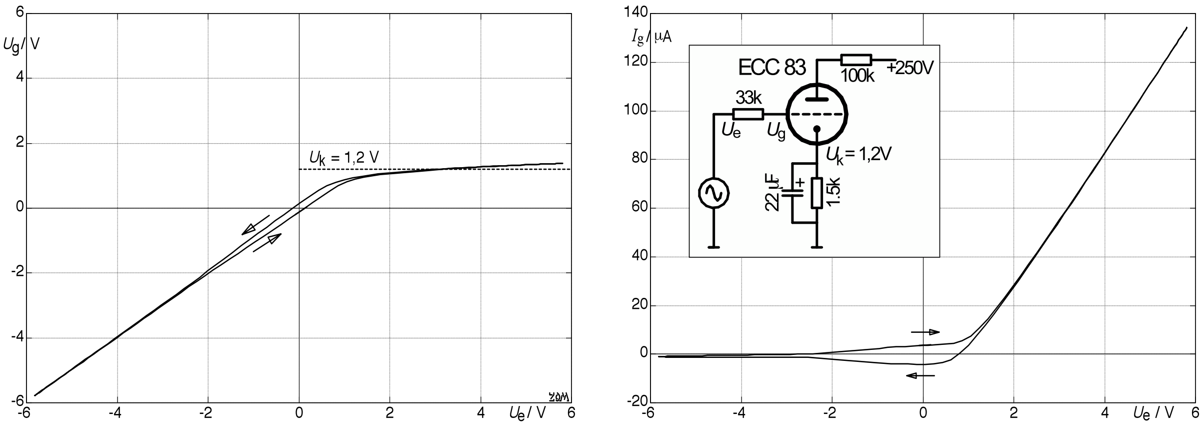

Fig. 10.1.2: Non-linear correspondence between input voltage U_e, grid-voltage U_g, and grid-current I_g.

Measurements of real tube voltages and tube currents show a hysteresis caused by capacitive coupling between plate and grid. Within the tube, the grid/plate-capacitance (about 1,6 pF) has an effect, and external stray-capacitances depending on the build of the circuitry weigh in. In conjunction with the grid-resistor, a low-pass in the feedback branch is created, i.e. the plate-voltage is (approximately) differentiated and the result superimposed onto the generator voltage. Since the plate-voltage is strongly limited for the drive signal shown in the figure (Chapter 10.1.3), this feedback becomes effective predominantly close to zero. Idealized characteristics are shown in **Fig. 10.1.3**: U_{gk} is the voltage between grid and cathode i.e. the actual control voltage of the tube. For the example it amounts to about -1,2 V in the operating point (i.e. without drive signal).

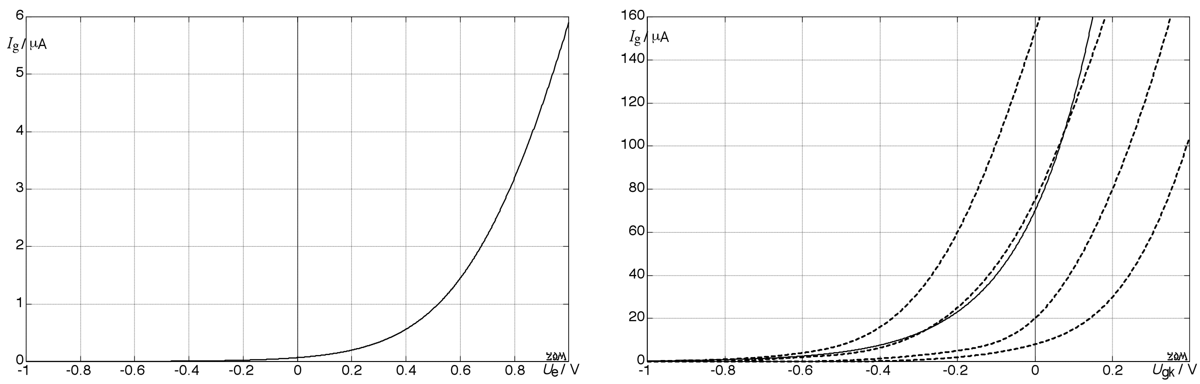

Fig. 10.1.3: Grid-current depending on the input voltage (or on the grid/cathode-voltage).
The right-hand picture shows measurements (----) in addition to the idealized curve.

♣ U_e between input and ground, U_g between grid and ground, U_{gk} between grid and cathode.

In order to assess the non-linear behavior, it is of course the magnitude to the actual input voltage that needs to be considered. If this were no larger than 100 mV, we could ignore the non-linearity. However, normal magnetic pickups can easily generate voltages in excess of 0,5 V, and even 4 V is not unheard of – therefore the non-linearity merits a discussion. The distribution function of the occurrence of pickup voltages is shown in **Fig. 10.1.4**. It may be nicely described by a Laplace-distribution (just like speech): the larger the amount of the pickup voltage, the less frequent it occurs♣. "Loud" pickups (Chapter 5.4), heavy strings and a strong picking attack may generate considerable voltages. The distribution function of this special example shows that 95% of all voltage values are smaller than 1 V, and 98% are smaller than 2 V. The relatively low likelihood of crossing these borders must not lead to the conclusion that the non-linearity may be neglected. Strong amplitudes especially happen with the plucking of a string (**Fig. 10.1.5**), and the immediate subsequent attack-process in the signal is analyzed by the hearing system with particular precision. The two signals shown in Fig. 10.1.5 do sound differently. However, the amplitude-limited signal – surprisingly – does not sound more distorted but less trebly than the original signal. The plate-voltage looks entirely different, again (Chapter 10.1.3), and what always holds is: the isolated portrayal of an individual non-linearity says little about the output signal of an amplifier.

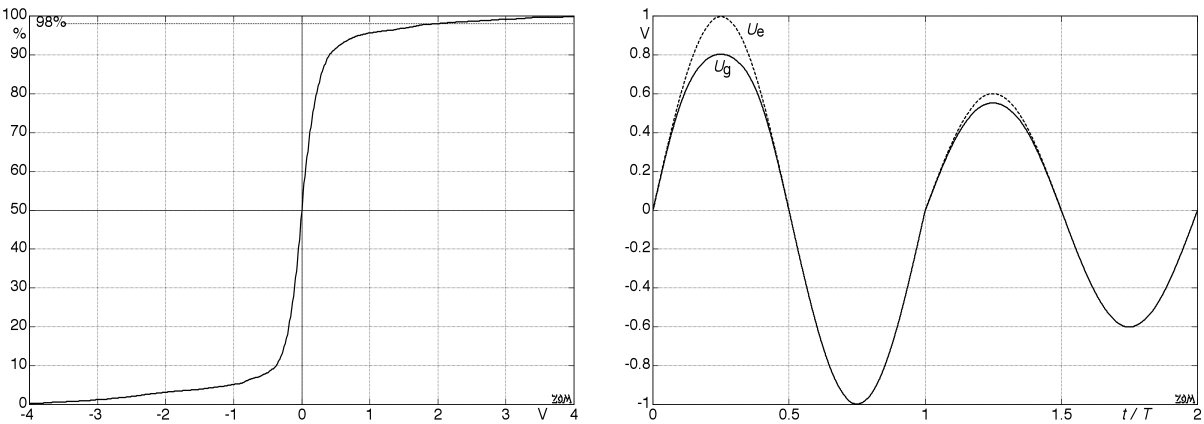

Fig. 10.1.4: left: Distribution function (cumulative) of the pickup voltage (Strat, SDS-1 in bridge position). Right: Non-linear correspondence between input voltage U_e and grid-voltage U_g for a sine signal.

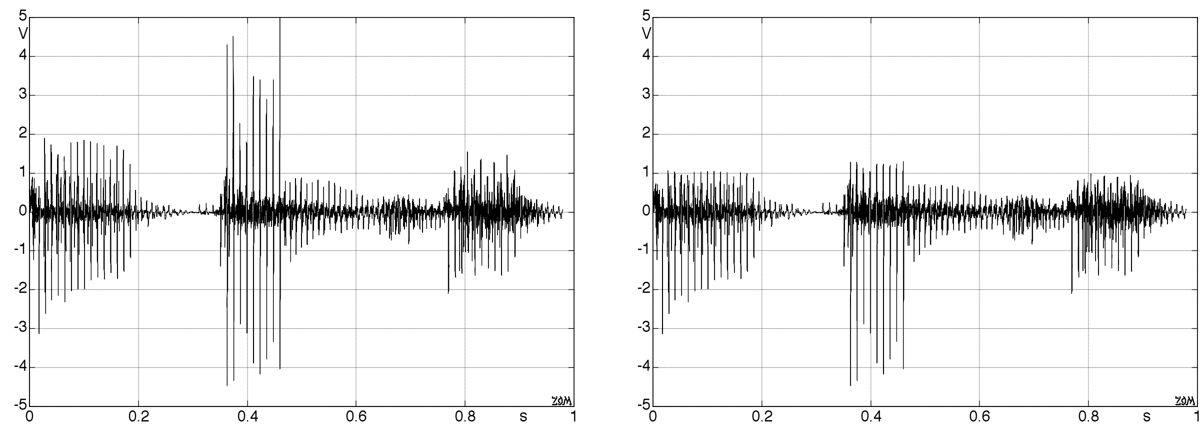

Fig. 10.1.5: Voltage-over-time at the terminals of an SDS-1 pickup (left); with limiting similar to a tube (right).

♣ Strictly speaking, the probability density is zero for <u>discrete</u> values of the continuously distributed voltage; to arrive at a probability (other than zero), integrating over a <u>range</u> is required.

10.1.3 Characteristic curves of the triode

The two circuits shown in Fig. 10.1.1 include a number of similarities; however, this must not lead to the conclusion that their behavior is equivalent. Already simple standard models indicate differences: for the tube, the correspondence between plate-current and control voltage is described via a power function with an exponent of 1,5 while for the FET, the corresponding exponent is 2. In reality, the characteristic curves of both amplifiers do deviate from this idealization – but not in the sense that they would become more equal.

Often, simple model-calculations for the triode start from the **Child-Langmuir-law**♣:

$$I_a = K_1 \cdot \left(U_a + \mu \cdot U_{gk}\right)^{3/2} = K_2 \cdot \left(U_{gk} + U_a/\mu\right)^{3/2} \qquad \text{Triode characteristic}$$

In this equation, U_{gk} designates the voltage between grid and cathode, and U_a designates the voltage between plate and cathode. K and μ are constants relating to the specific tube while I_a is the plate-current. As simple as this law is: it is as inappropriate for guitar amplifiers. Differences relative the real triode already show up in the range of the characteristic curve that could be seen as reasonably linear; for the overdrive range, the Child-Langmuir-law utterly fails (it was not put together for this scenario, anyway). **Fig. 10.1.6** compares idealized and real triode characteristics – the differences are significant. In literature (e.g. JAES), we find several improvements of the above equation that brings it closer to reality (i.e. closer to the characteristics given in data books), but the resulting complex equations do not only require two but six or even more modeling parameters. If the latter are optimized to model the linear and the weakly non-linear drive range, we may still not assume that the extreme overdrive conditions♥ in guitar amplifiers are also suitably modeled. The following depictions therefore do not orient themselves according to tube models but are based on actual precision-instrumentation-measurements taken from amplifier-typical circuits. This included all the associated uncertainties … whether this exact circuit or this tube-specimen was typical enough, whether the capacitors had been run-in long enough, whether the moon had already risen (or set, or was in the correct house) …

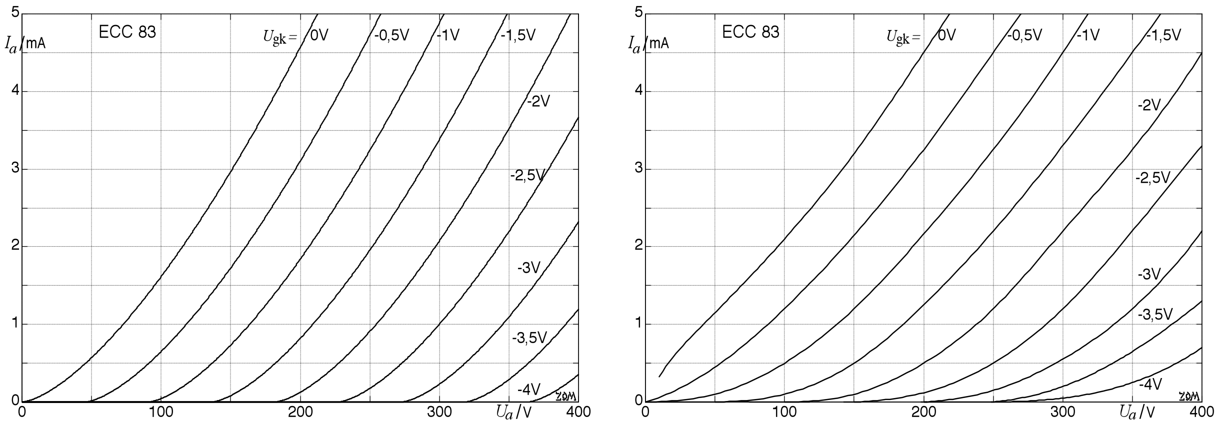

Abb. 10.1.6: Tube characteristics. Left: idealized according to Child-Langmuir. Right: data sheet info.

♣ D. Child: Phys. Rev., Vol. 32 (1911), p.498. I. Langmuir: Phys. Rev., Vol. 2 (1913), p.450.
♥ The drive-limit for the linear range may easily be exceeded by a factor of 30.

The plate-current I_a depends on both the control voltage♣ U_{gk}, and on the plate-voltage U_a (Fig. 10.1.6). In a real amplifier circuit (Fig. 10.1.1), both of these values change, and consequently the transmission behavior may not be taken as such from Fig. 10.1.6. **Fig. 10.1.7** therefore directly indicates the mapping of the input voltage U_e onto the plate-voltage U_a. In the left part, the input voltage (multiplied by a factor of -58) is included, as well, in order to clearly show the effect of the non-linearity: both half-waves experience limiting. The latter is explained only from the overall transmission behavior, and not merely from the input characteristic. The right hand part of the figure shows the plate-voltage for input voltages of 1 V_{eff} and 4 V_{eff} respectively.

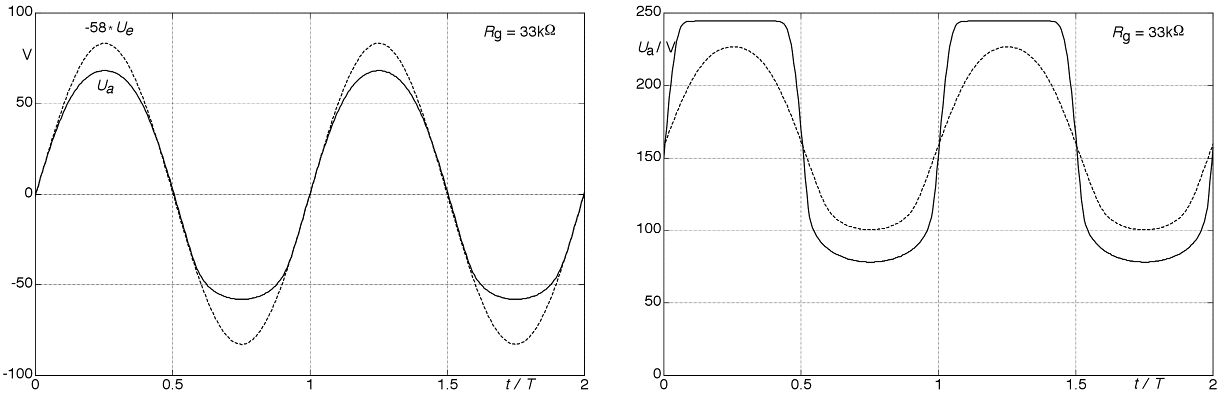

Fig. 10.1.7: ECC83: non-linear distortion of the plate-voltage; on the left with vertical offset. R_a = 100kΩ.

The figure shows how the negative half-wave is flattened first as the drive level increases; for strong overdrive, heavy **clipping** is introduced for the positive half-wave. The plate loading (5 MΩ-probe) is the reason why the plate-voltage does not fully reach the supply-voltage (250 V). That the minimum voltage is not closer to zero is due to the grid-resistor – it attenuates positive input voltages on their way to the grid (Fig. 10.1.4) and prohibits full drive of the tube. **Fig. 10.1.8** shows the influence of the grid-resistor: without R_g, larger plate-currents and smaller plate-voltages are possible – this kind of operation is, however, not typical for input stages of customary guitar amplifiers, and it will not be investigated further. What does require consideration is the **plate-load** that has, in the measurements so far, been very small (at 5 MΩ). In the classic tube amps (Fender, VOX, Marshall), the input tube often feeds the **tone-control stage** that exerts considerable loading onto the plate.

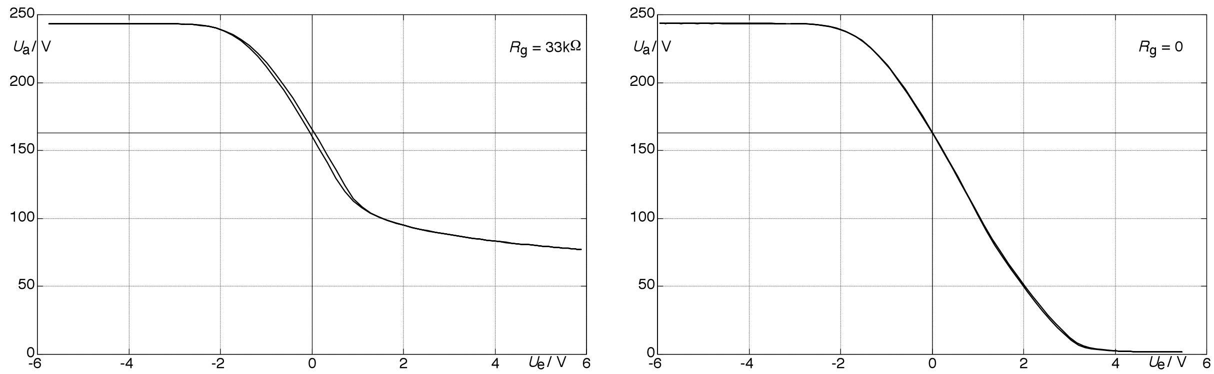

Fig. 10.1.8: transmission characteristic $U_e \rightarrow U_a$. For a 33-kΩ-grid-resistor (left); for shorted grid-resistor (right). R_a = 100 kΩ (plus 5 MΩ load).

♣ May take different meanings; in this case: grid/cathode-voltage.

The input impedance of the tone-control stage is complex, and therefore the analytical description now begins to become complicated (non-linear and frequency-dependent behavior). As a first approximation, however, we may replace the input impedance of a typical tone-control network by the series connection of a 100-kΩ-resistor and a 0,1-µF-capacitor – this enables us to describe the important effects already pretty well. More precise, amplifier-specific models would go beyond the scope of these basic considerations. The cutoff-frequency of the load-two-pole is low enough that the plate is loaded with 100 kΩ in the steady-state condition. Compared to the situation without load (as it has been looked at so far), the AC-plate-voltage is reduced by about a third (**Fig. 10.1.9**). The measured **small-signal-gain**, i.e. the gain for small drive levels (e.g. 0,1 V), amounts to -42. In theory, the small-signal-gain results from the multiplication of the transconductance S (data sheet: S = 1,6 mA/V) with the operational resistance. The latter consists of the parallel-connection of the internal impedance of the tube (data sheet: 63 kΩ), the plate-resistance (in the present example 100 kΩ), and the load resistance (again 100 kΩ). We calculate a small-signal-gain of -45, from this i.e. a reasonable correspondence. What needs to be borne in mind, though: the data-sheet information may be taken only as a guide number: swapping a tube for another can easily change the small-signal gain by 3 dB! The drive limits are specific to the respective tube specimen, as well.

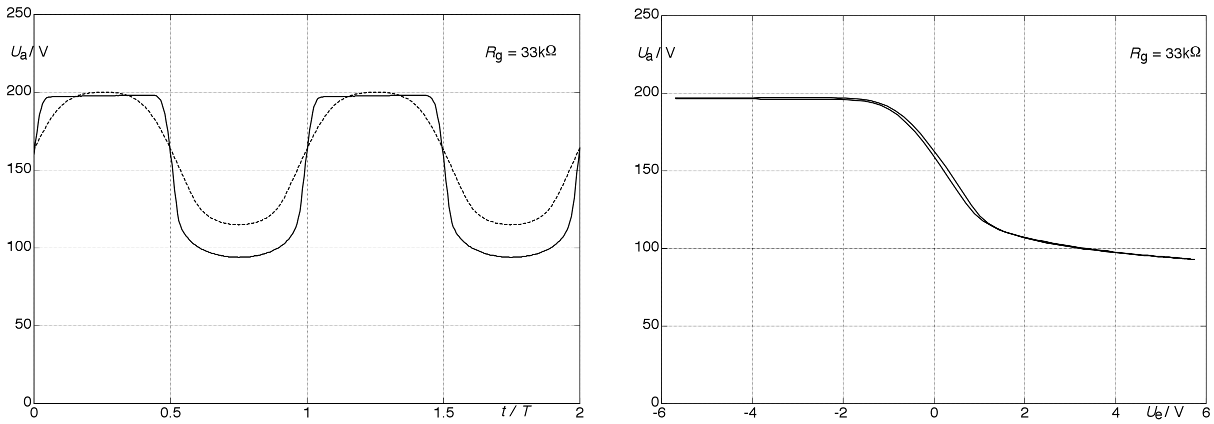

Fig. 10.1.9: Plate-voltage for input voltages of $1V_{eff}$ and $4V_{eff}$ (with R_g, and plate loading, left); transmission characteristic (with grid-resistor R_g and plate loading, right).

The comparison between Figs. 10.1.7 and 10.1.8 has already shown how important the internal impedance of the signal generator is. Whether the tube grid is driven from a low-impedance source (R_g = 0), or via a grid-resistor (R_g = 33 kΩ) makes a big difference. Of course, the serially connected impedance of the signal generator needs to be considered in addition. Active pickups (e.g. EMG) feature internal impedances similarly low as those of the generators used for the measurements; however, most guitars have high-impedance passive pickups. For an exact analysis, the operation with a 50-Ω-generator is therefore not indicative of the behavior when driven by an electric guitar. The latter may easily show an internal impedance of 100 kΩ in the range of the pickup resonance (2 – 5 kHz). Since the **internal impedance** of the electric guitar is frequency dependent (e.g. 6 kΩ at low frequencies and 100 – 200 kΩ at resonance), and since the input impedance of the tube is non-linear, complicated interactions between the different systems occur already in the input stage of a tube amp. Such an amp will make the guitar see an entirely different load compared to a “modeling amp”. In the latter, the guitar-signal will be normally fed - via a high-impedance OP-amp-stage - to the AD-converter, and all signal processing will be taken care of in the digital realm. However, which tube characteristics will in the end lead to audible differences can only be investigated via listening-experiments.

For two different source impedances, the grid-voltage limiting is shown in **Fig. 10.1.10**. It is evident that even relatively small positive voltages are visibly reduced. The internal impedance of the generator is, however, purely ohmic in this measurement – which does not correspond to the situation for a connected electric guitar fitted with passive pickups. To better simulate this operational condition, a small **transmitter coil** was laid on top of the pickup of a Stratocaster (with original wiring). Driving this coil with a power amplifier generated a magnetic AC-field inducing a sinusoidal voltage into the pickup. The internal impedance of this arrangement therefore realistically corresponded to the actual operation.

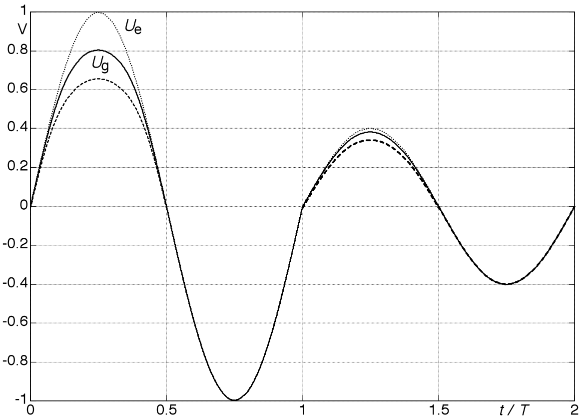

Fig. 10.1.10: Limiting of the generator voltage (U_e) in the grid circuit. Internal generator impedance = 0 (——), and 100 kΩ (----), plus grid-resistor R_g = 33 kΩ. Two periods with different voltage-amplitude shown (1V and 0,4V).

In its left-hand section, **Fig. 10.1.11** shows the corresponding measurement results. The dashed line relates to the source voltage of the guitar corresponding to the open-loop voltage generated by the unloaded guitar. With the load of the tube amplifier, the guitar voltage is bent out of shape; however, this does not happen such that the positive half-wave would simply be compressed (as it would be the case for an ohmic source impedance). Rather, the complex guitar impedance leads to phase shifts between the spectral distortion components (especially in the 1st and 2nd harmonic), and thus the voltage curve is also changed for the negative half-wave. The grid-voltage changes correspondingly (right hand section of the figure), and in the plate-voltage the duty factor is shifted (compare with Fig. 10.1.9). These measurements show that already the first interface between guitar and amplifier-tube has an effect on the signal. Precise observation indicates that the tube input is not of ideally high impedance but acts as a non-linear load-resistance already at moderate voltages. Whether the corresponding changes in the signal are audible compared to other non-linearities, is another question and can, however, be determined only for the individual case.

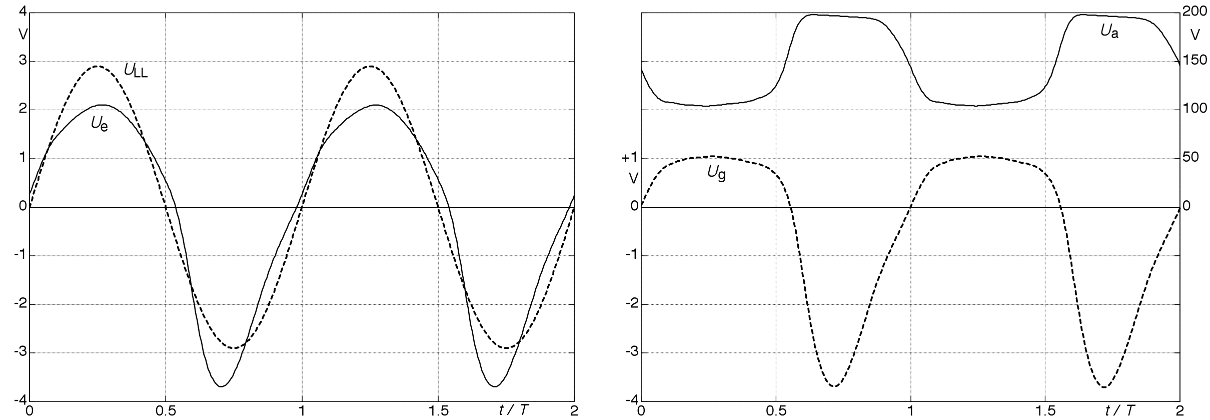

Fig. 10.1.11: Mapping of the guitar-source-voltage U_{LL} onto the terminal voltage of the guitar (left). Load resistance for the guitar is the input-circuit of the tube, R_g = 33 kΩ, f = 2kHz. On the right, the corresponding grid- and plate-voltages are shown; the plate is loaded as given in Fig. 10.1.9.

It is almost impossible to describe the transmission behavior of a guitar amplifier in its entirety by formulae and diagrams. This is not because the relations and connections would be unknown, but rather because too many dependencies would have to be defined. While the small-signal behavior can easily by specified via the frequency-response, there is – strictly speaking – not even a transfer-function for the large-signal operation because this function is only defined via the LTI-(linear time-invariant)-model. Mixtures of small-signal frequency-response and harmonic-distortion characteristic are either incomplete or too extensive. Non-linear distortion is dependent on frequency and on level, and thus is a bi-variant quantity. There are, in fact, *many* bi-variant quantities: 2nd-order (k_2) and 3rd-order (k_3) harmonic distortion, as well as 2nd-order and 3rd-order difference-tone-distortion, just to name the most important ones. For the – frequently occurring – strong-overdrive condition it is not adequate in the least to assess distortion up to merely the 3rd-order; rather, it would be necessary to determine a multitude of individual harmonic-distortion- and difference-tone-factors, and represent this as a function of two variables. And even if we would make such an effort: the result would be all but impossible to interpret. For example, how would we evaluate a circuit change that results in a reduction of the 3rd-order harmonic distortion at 0,5 V and 1 kHz, while the 2nd-order harmonic distortion increases at 0,8 V and 2 kHz? While at the same time the 4th-order harmonic distortion at 0,8 V (2 kHz) drops strongly but the 2nd-order difference-tone-distortion generally grows stronger? Is this desirable or counter-productive? General judgments such as: *for tubes, 3rd-order distortion (= good) dominates, for transistors 2nd-order (=bad) does* are far too unsophisticated, but unfortunately they keep getting copied again and again from textbook to textbook. Listening experiments remain indispensable. Still, a few fundamental relations can be taken from the theoretical models, after all – even if it the result is not much more than the insight that the circuit layout (than cannot be derived from the schematic) can be highly important, or that tube data have a considerable scatter range. In the following analyses, we will give some data on harmonic distortion for a tube driven via an ohmic source impedance, all the while remaining fully aware that only part of the topic can be covered this way, and that additional research would be highly desirable.

10.1.4 Non-linearity, harmonic-distortion factor

Here is a simple **example** regarding the topic of non-linearity: an amplifier generates – at an input voltage of 1 V – pure 2nd-order harmonic distortion with k_2 = 5%. Let's set its gain factor to v = 1. Now, a second amplifier (also with v = 1) also generating k_2 = 5% at 1 V is connected in series with the first one in a non-reactive fashion. How big would the harmonic distortion of the overall system be?

Would that be: k_2 = 10%, or 7%, or an unchanged 5%?

It is not even possible to answer this question without supplementary data: we do not know the phase of the distortion. In case the 2nd-order distortion is generated in-phase in both amplifiers, k_2 is doubled, but if it happens to be in the opposite phase, the 2nd-order distortions all but cancel themselves out. In both cases an additional 3rd-order distortion appears at k_3 = 0,5%. If there is a random phase-shift between the two amps, k_2 can assume any value between 0 and 10%. Already this simple example shows that it is very difficult to derive any statements about the distortion of the overall system from the non-linear behavior of the single amplifier stages.

So, are you having fun yet, dear audio-engineers? O.K. then – let's go for a **second example**: now both amplifier stages feature pure 3rd-order distortion at k_3 = 5%. Right … use the above: the series connection results in k_3 = 10% for the in-phase condition, and for the out-of-phase condition in k_3 = 0%; plus additionally k_4. Hm … are you sure? Then do turn the page!

For the pure 3rd-order distortion, the overall system does not distort with k_3 = 10%, but with k_3 = 12,3%, and k_5 = 1% is generated in addition, rather than k_4. Given the anti-phase-condition, we do not see a cancellation but k_3 = 7,5%! Even examples as simple as these show that the results of connections of non-linear systems are rarely understood based on intuition. Moreover, tubes do not exhibit *pure* 2nd-order distortion or *pure* 3rd-order distortion; there will also be distortion of higher order, and in addition the signal will be subject to filter stages – with the result being a highly complex signal processing despite the relatively simple circuitry.

Often, modeling a non-linear circuit starts with the simplification that the system is memory-free. With the investigated system not including any signal memory, the output signal exclusively depends on the input signal at the same instant – with the dependency between both signals described by the **transmission characteristic** (not the transmission function!). This transmission characteristic $y(x)$ is curved (compare to Fig. 10.1.8), but it is time-invariant and excludes any hysteresis. The characteristic may be expanded into a power-series (Taylor/MacLaurin) around the operating point – the smaller the drive levels, the more precisely this works. Put in another way: the more the amp is driven, the less the power series is appropriate. This is easily understood: a limiting characteristic has two horizontal asymptotes, which is incompatible with a power-series converging towards infinity. In this situation, wouldn't the arctan-function seem to be a much better starting point? Yes, indeed – but it would be one that is far from intuitively accessible: how is e.g. $x = \hat{x} \cdot \sin \omega t$ mapped again onto y? With $y = \arctan(\hat{x} \cdot \sin \omega t)$. O.k., I see. So how does the harmonic distortion depend on the drive levels? Well, we would have to develop a series-expansion of the arctan … Phewww – that means we might as well expand the transmission characteristic into a series:

$$y = a_0 + a_1 \cdot x + a_2 \cdot x^2 + a_3 \cdot x^3 + ... \qquad \text{Series expansion of the transmission characteristic}$$

In this expansion, a_0 is the DC-offset that is separated in most circuits by the coupling capacitors; we ignore this offset. a_1 is the gain – for an input tube this might be e.g. -54. Now we get to the non-linearity: using, for example, the pure 2nd-order distortion (i.e. $a_i = 0$ for $i > 2$), we obtain

$$y = a_0 + a_1 \cdot \sin \omega t + a_2 \cdot (\sin \omega t)^2 = a_0 + a_1 \cdot \sin \omega t + a_2 \cdot (1 - \cos 2\omega t)/2$$

Due to the non-linearity, the DC-component has changed but we can again ignore it. There is now a new spectral line at twice the frequency. The ratio of the RMS-values $k_2 = \tilde{y}_2 / \tilde{y}$ is designated the **2nd-order harmonic distortion** k_2. $\tilde{y}_2$ is the RMS-value of the 2nd-order harmonic (at 2ω) and $\tilde{y}$ is the RMS-value of y. Let us set, as an example, a_1 = 1 and a_2 = 0.1 – this yields $k_2 \approx a_2 / 2a_1 = 5\%$. Connecting two such systems in series, a series-transformation $z(y(x))$ is the result:

$$z = a_0 + a_1 \cdot y + a_2 \cdot y^2 = a_0 + a_1\left(a_0 + a_1 \cdot x + a_2 \cdot x^2\right) + a_2\left(a_0 + a_1 \cdot x + a_2 \cdot x^2\right)^2$$

Assuming again $x = \sin(\omega t)$, the amplitudes (or rather the RMS-values) of the individual harmonics can be calculated. What is striking in view of the second bracket of the equation is that the offset (a_0) now not only influences the DC-component but the 1st and 2nd order harmonic, as well! Moreover, we notice the generation of a 4th-order harmonic due to x^4 – although its amplitude is so small that it may be disregarded. From x^2, a DC-component and the 2nd-order harmonic result, and from x^3 we derive the 1st-order and the 3rd-order harmonics. x^4 generates a DC-component plus the 2nd-order and 4th-order harmonics. So, everything depends on everything else, more or less.

In summary: for $a_0 = 0$, the levels of the 1st, 2nd, and 3rd harmonic depend on a_1 and a_2, and only the 4th harmonic depends solely on a_2. A simplification yields:

$$z = x + 2a_2 \cdot x^2 + 2a_2^2 \cdot x^3 + a_2^3 \cdot x^4 ; \qquad \text{for } a_0 = 0 \text{ and } a_1 = 1.$$

Neglecting higher-order effects, we can state that the series connection indeed leads to double the 2nd-order harmonic distortion. Also, if one system is set to be $y = a_1 \cdot x + a_2 \cdot x^2$ and the other is set to be $z = a_1 \cdot y - a_2 \cdot y^2$, k_2 can be approximately fully compensated. The simplifications we introduce here are purposeful – they do not generate any large errors.

For the purely 3rd-order distortion, the mapping is: $y = x + a_3 \cdot x^3$. The offset a_0 is set to zero in this case, and the linear term is set to 1 ($a_1 = 1$) as a simplification. The series connection yields:

$$z = y + a_3 \cdot y^3 = x + a_3 \cdot x^3 + a_3 \cdot \left(x + a_3 \cdot x^3\right)^3 = x + 2a_3 \cdot x^3 + 3a_3^2 \cdot x^5 + 3a_3^3 \cdot x^7 + a_3^4 \cdot x^9$$

Again, we could disregard all higher-order terms and assume that the 3rd-order harmonic distortion will be doubled. However, the 1st and the 2nd harmonic are dependent on all summands, and the resulting effect is not at all that minor:

$$\sin^3(\varphi) = \frac{1}{4}\big(3\sin(\varphi) - \sin(3\varphi)\big), \qquad \sin^5(\varphi) = \frac{1}{16}\big(10\sin(\varphi) - 5\sin(3\varphi) + \sin(5\varphi)\big)$$

The summation of all terms of the expansion has the effect that the RMS-value of the 3rd harmonic is not only doubled but rises by a factor of 3,7. At the same time, the RMS-value of the overall signal increases by half, yielding k_3 = 12,3%. If the sign of a^3 in one of the two systems is inverted, x^3 can be reduced to zero but the remaining members of the series deliver a significant contribution to the 3rd harmonic. The amplitude of the latter therefore does not go down to zero but decreases merely to 7,5%.

If the **offset** (a_0) is not set to zero, the situation becomes even more complicated. The same happens if we do not keep the limitation on purely 2nd- and 3rd-order distortion, respectively. So: with two nonlinear systems connected in series, *both* generate 2nd- as well as 3rd-order distortion.

$$z = b_0 + b_1 \cdot y + b_2 \cdot y^2 + b_3 \cdot y^3 ; \qquad y = a_0 + a_1 \cdot x + a_2 \cdot x^2 + a_3 \cdot x^3$$

O.k. – computing this is not impossible; we could multiply that out. The added value would not be that big, however. It is already clear now that the RMS-value of each harmonic will be dependent on many coefficients. Also, there will be cancellations of components if there are opposing algebraic signs. These cancellations will be drive-level-dependent, though – or at least there will be drive-level-dependent maxima and minima. The individual harmonic distortion components will not simply experience a monotonous increase with rising drive-levels but can pass through complicated curves. The individual system may generate exclusively even-order distortion (k_2, k_4, ...), but the series connection of two such systems may still show a predominant odd-order distortion. If we now consider that even simple guitar amplifiers do not contain one but four tube-stages, and that each tube introduces distortion both at its input *and* at its output, and that moreover tone-filters will change amplitudes and phases … this is where we start to catch a glimpse of how complex a guitar amp in fact is.

The following figures present the dependency of individual harmonic distortion components on the input signal level. N.B.: Another way of quantifying distortion is expressing how much lower the level of the distortion is compared to the original (undistorted) signal. This approach yields the so-call harmonic **distortion attenuation** a_{ki} calculated from the harmonic distortion factor k_i (as we have been using it so far) as:

$$a_{ki} = 20 \cdot \lg(1/k_i)\,\mathrm{dB} \qquad \text{for e.g. } k_2 = 5\% \;\rightarrow\; a_{k2} = 26\,\mathrm{dB}.$$

All measurements were done with a regularly heated ECC83 with the intrinsic distortion factor of the analyzer being negligible (k < 0,001%, CORTEX CF-100). The cathode of the tube was connected to ground via 1,5 kΩ // 25 µF (Fig. 10.1.1), and the plate to U_B via 100 kΩ. To model the load, the plate was additionally connected to ground via a 0,33-µF-100-kΩ-series-circuit. The signal was fed to the grid from a low-impedance generator (CORTEX CF-90) via the grid-resistor R_g. For one row of measurements, R_g was 33 kΩ (corresponding to a classic tube amp scenario that is fed from a low impedance source), and for the other row it was R_g = 133 kΩ (corresponding the additional source impedance of 100 kΩ as it can be present if a guitar with a passive pickup is operated around its resonance frequency, compare to Fig. 10.1.10). The supply-voltage U_B amounted to 200 V and 250 V, respectively, i.e. typical settings for input stages (**Fig. 10.1.12**).

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Abb. 10.1.12: Distortion attenuation as function of the generator level, R_g and U_B vary. 0 dBV ≙ 1V_{eff}.
These graphs are reserved for the printed version of this book.

While the a_{k3}–curve maintains its shape and predominantly experiences “merely” a shift, the minima and maxima of the 2nd-order distortion change rather drastically. *So will you tell me how that sounds, already?* would be an obvious question … however: nobody actually listens to the plate-voltage of the preamp-tube, and therefore the *sound* of that signal is irrelevant. Highly relevant would be how the differences mentioned above affect the loudspeaker voltage, but this would require the consideration of a myriad of additional parameters and go beyond the constraints give here. Unfortunately.

Another question relates to the **tube**: RCA, Tungsram, Telefunken, Chinese, Russian, NOS, little/much used, and whatever other difference there might be? Simple answer: the tube came out of the box that served here as container for tubes since 1965, and was re-stocked many times since. An ECC83 cost DM 7,50 (about € 3,25) in Germany in 1965; today is offered for € 6.-. It could also set you back € 25, or even more, though. Without a doubt, tubes of the same type can differ a lot – the label "ECC83" does not indicate any special sound. Selection processes performed by the supplier *may* be helpful but do not *have* to be. Pricy tubes are not necessarily in principle better than cheap ones; in particular, "NOS" (i.e. the tube that has spent 50 years on the shelf without being touched) does not guarantee a "super-sound".

In **Fig. 10.1.13** we see differences that can occur when we change tubes (all measurements taken with ECC83s). A tube was simply unplugged and another was plugged in, instead. It is intentional that manufacturers are not identified here, since we do not have a representative sample. We did not investigate whether an old 80-$-NOS-tube delivers similar or entirely different curves – confronted with its measurements, it might have experiences a kind of final deadly shock. Plus, strictly off the record: for the analyst, this is somewhat like the situation experienced by Galileo's colleagues who did not even want to look through the telescope to see Jupiter's moons – some of us in fact don't really want to know.

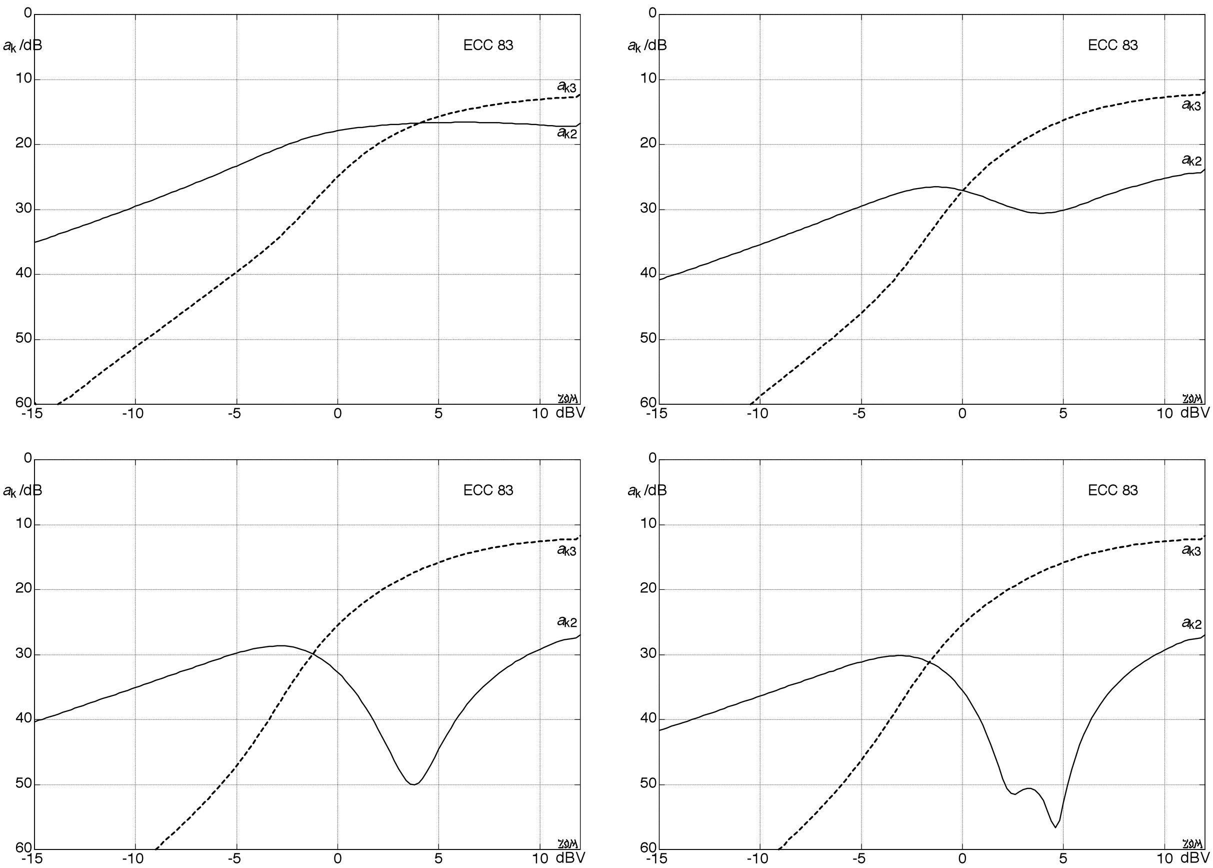


Fig. 10.1.13: Differences in harmonic distortion attenuation caused by swapping tubes. U_B = 250V, R_g = 33kΩ.

Now back *for* the record: already at an input voltage of 300 mV_{eff}, the harmonic distortion in the input stage of a guitar amp can reach 3%. For small input voltages, 2^{nd}-order distortion is predominant while from 0,25...1 V, 3^{rd}-order distortion dominates. The location of the border between the two distortion types depends on the grid-resistor, on the supply-voltage, and on the ECC83-specimen. The distortion is not inherently unwelcome but rather typical for a guitar amp of this construction.

So far, we have varied, as parameters, the grid-resistor, the supply-voltage and the tube itself. As the **supply-voltage** changes, the plate-current, the plate-voltage, and the grid/cathode-voltage change, as well. Of course, more parameters vary – but right now we look only at these three. For example: increasing the supply-voltage from 200 V to 250 V increases the plate-voltage from 131 V to 165 V, and U_{gk} decreases from -0,97 V to -1,23 V. Another method to vary the anode current is the so-called "cathode clamp": here, the cathode-voltage is imprinted (i.e. kept constant) using a separate power supply. One could think that the cathode-voltage could not change anyway due to the capacitor connected in parallel – but in fact, it can: a 2nd-order-distorted sine tone will generate a DC-component ($f = 0$) that shifts the operating point. The following figures show the effects of a relatively small change in the grid/cathode-voltage on the distortion /(**Fig. 10.1.14**).

Abb. 10.1.14: Harmonic distortion-attenuation dependent on the input level with varying grid/cathode-voltage. These figures are reserved for the printed version of this book.

It is clearly visible that even apparently minor changes in the operating point have considerable effects on the non-linear distortion. For reasons of clarity, no higher-order distortion products are included in the figures; it can be stated, however, that they are highly similar. The operating point of the tube is far from fixed but drifts while the amp is being played. One cause for this is found in the non-linearities already mentioned, and another lies with the time-variant supply-voltage. The latter depends on the plate-currents of the power stage and the internal impedance of the power supply and will change depending on the output power of the amp at the given moment (see Chapter 10.1.6). For the Fender Deluxe we investigated, this variation was as much as between 210 and 247 V, after all . . .

Now, what is so special in a tube amplifier compared to other amps? Looking at the preamplifier, there are differences in particular in the non-linear behavior. There are, in addition, compressor effects and linear filtering – this will be elaborated upon a bit later. The **operational amplifier** (OP) appears to be a modern alternative to the tube. It has an operational range to above 1 MHz and its harmonic distortion may be reduced to 0,001%. These are, however, all properties that a guitar amplifier should not actually have! An OP may only be considered as an alternative if additional circuitry simulates the non-linear behavior of the tube. That this is not entirely trivial was shown in the preceding paragraphs. **Fig. 10.1.15** depicts the drive-level-dependent increase of the distortion for the ideal OP in comparison with to the tube. The hard amplitude limiting ("clipping") leads to a steep distortion increase that is atypical for a tube.

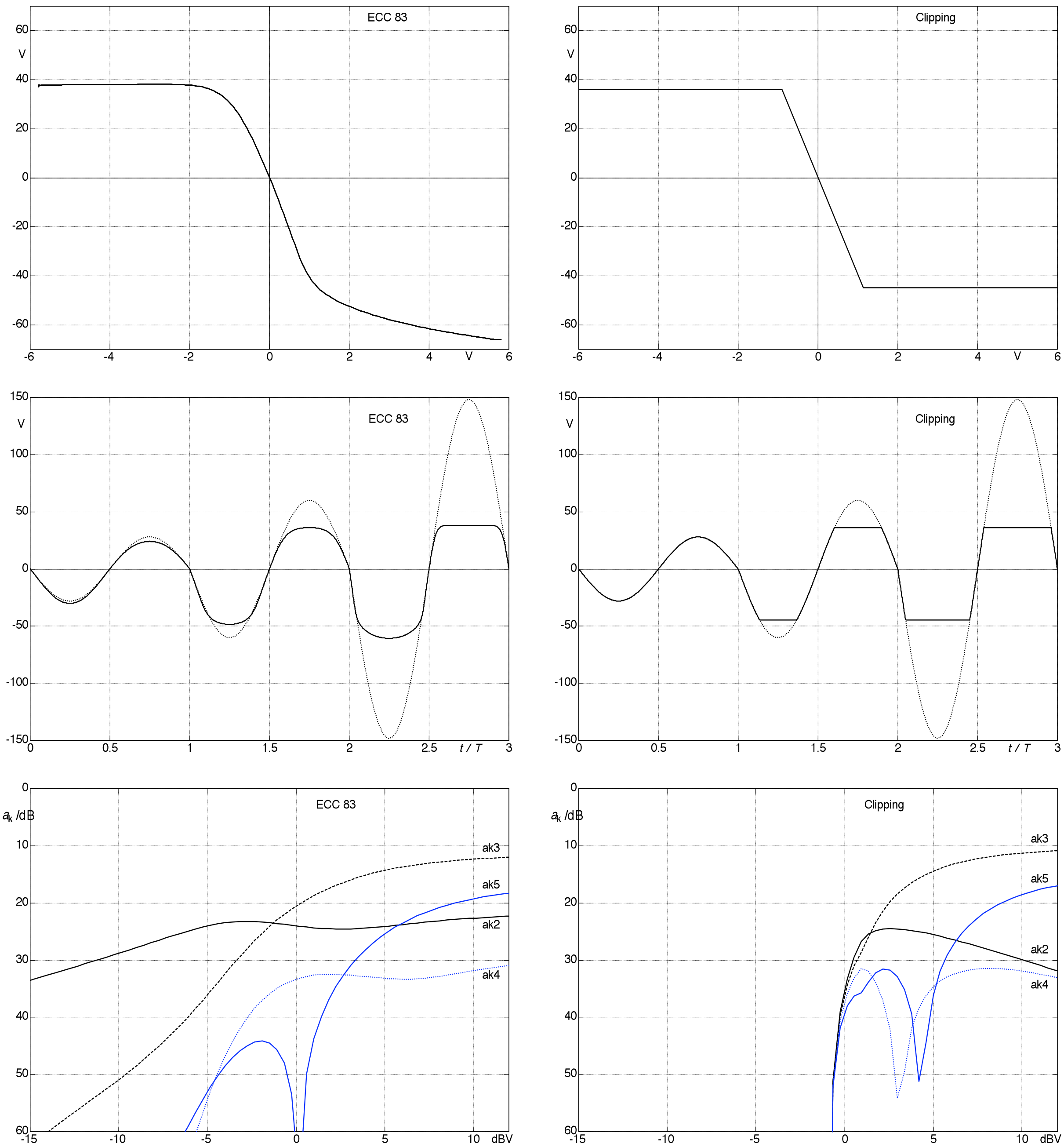


Fig. 10.1.15: Distortion for tube-typical limiting (left) and hard OP-clipping (right). By the way: the designation "ideal OP" does not imply that the OP would be ideal for playing guitar through it. NB: The OP-offset was adjusted for an asymmetry similar to a tube.

The rise of the distortion will be considerably flatter if the signal limiting is realized not by the OP itself but by two silicon diodes (1N4148) in an anti-parallel connection (**Fig. 10.1.16**). If the two measured diodes were perfectly identical, only odd-numbered distortion products would appear; due to small production-related differences, we also obtain even-numbered distortion products in this example. The 3rd-order distortion of this diode circuit already shows strong similarities to the triode circuits measured in Fig. 10.1.15 but the 2nd-order distortion is not reproduced yet. It is not very demanding to design – using a combination of germanium and silicon diodes – a non-linear two-pole the distortion behavior of which sounds similar to that of a tube. The exact reproduction of tube-distortion is not even required for this; an approximate modeling suffices.

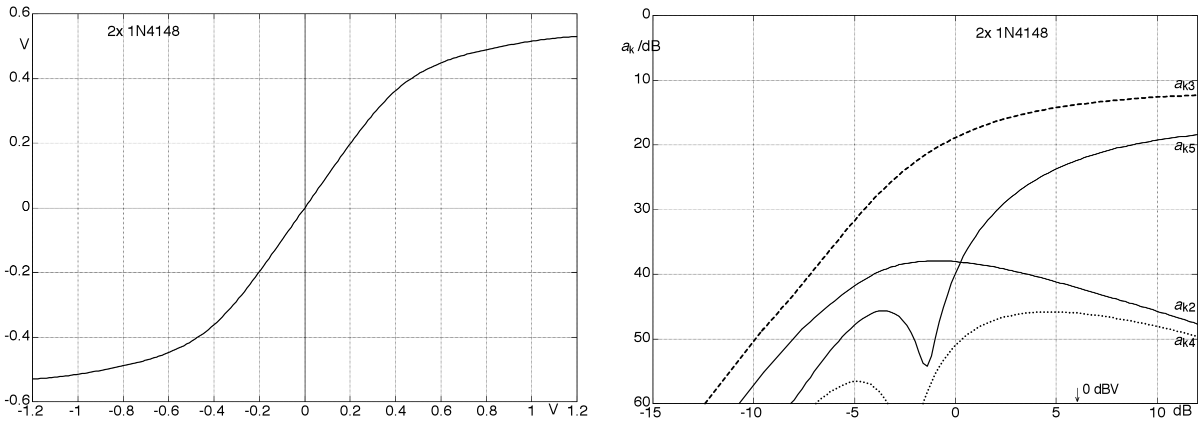


Fig. 10.1.16: Signal limiting using two anti-parallel silicon diodes (1N4148) fed from a 20-kΩ-resistor. The level reference on the abscissa of the right-hand picture is chosen to match the representation in Fig. 10.1.15.

It is not only the harmonic distortion that is different in tube and OP, but the **compression** is, as well (**Fig. 10.1.17**). This difference is not big, but may be compared to the so-called "sagging" – a modulation caused by the power supply (Chapter 10.1.6). In the attack phase of a tone, a tube amp may lend that extra little bit of power that can be decisive when competing with other instruments. That tube amplifiers can be louder than transistor amps rated at the same power is due in particular to the higher output impedance, but may also have to do with the weaker compression (= increased dynamics). Of course, this is not generated by the preamp-tube alone but by the overall circuit.

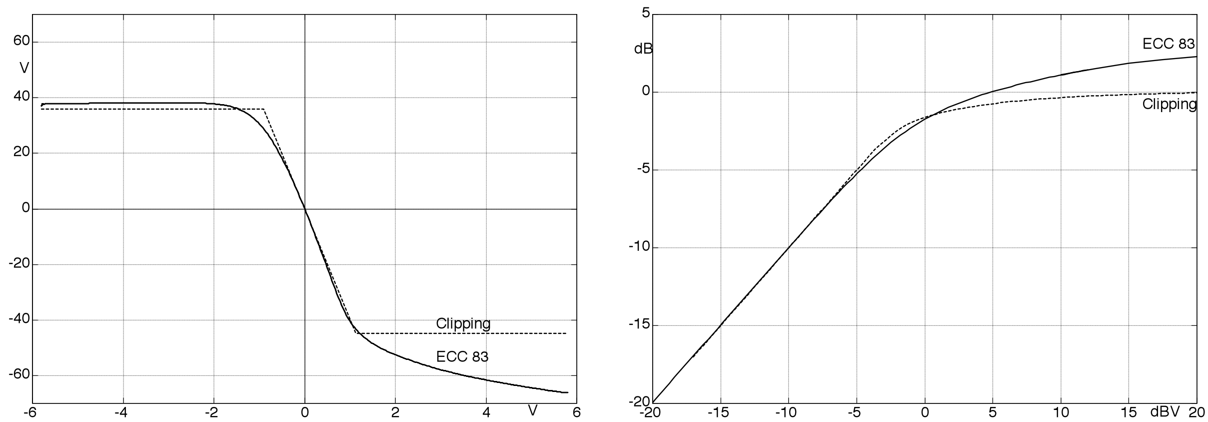


Fig. 10.1.17: Signal-limiting in a tube and an ideal OP. Equal small-signal gain.

We will dedicate a later chapter to the power-delivery, but at this point a fundamental aspect of the connection of non-linear systems may already be briefly introduced: the transformation caused by the individual systems are generally not **commutative**, i.e. the individual systems cannot be simply interchanged in their sequence. For this reason, it is not possible to replace an amplifier consisting of a plurality of non-linear and linear systems by a single non-linear stage and a single filter-stage. Special consideration needs to be given to the fact that already the coupling capacitor that taps the signal from the plate is such a filter-system, even if the associated cutoff frequency of this high-pass is very low.

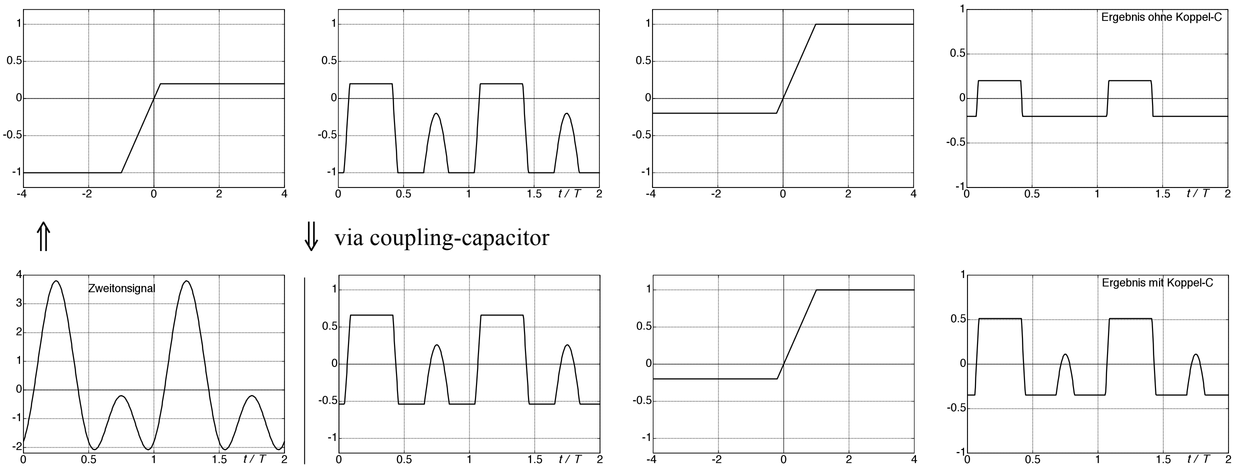
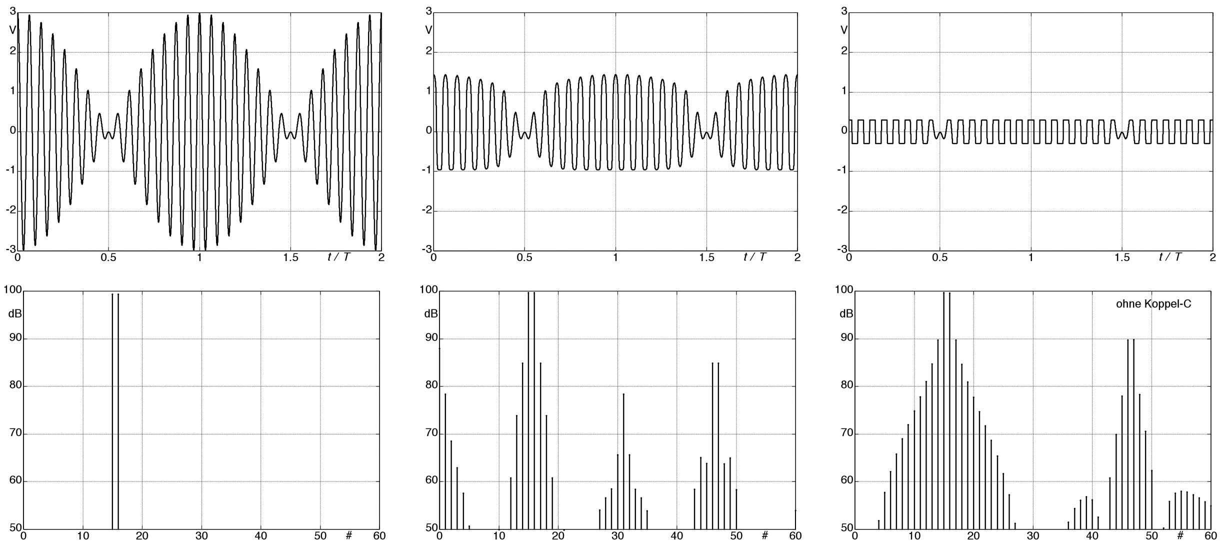


Fig. 10.1.18: Half-wave limiting in a sequence of systems; top: w/out coupling cap, bottom: with coupling cap.

Fig. 10.1.18 shows an example: a two-tone signal first passes a stage limiting the positive half-wave and then a second stage limiting the negative half wave. If these two stages directly follow each other, the result is a signal limiting on both sides as depicted in the upper row of pictures. However, if a **coupling capacitor** is connected in between the two limiting stages, we obtain an entirely different output signal (lower right picture). With the coupling capacitor connected *ahead of* the first limiter stage, it would have no effect because the two-tone signal is already without any DC-component. The same result would be obtained with the capacitor positioned *after* the two limiting stages. However, connected *between* the two stages, the capacitor will change the signal even if the cutoff frequency is far below the two frequencies contained in the two-tone signal. Now, let's put this in the context of a guitar amplifier: since the plate-voltage is 150 – 200 V even without any drive signal, a coupling capacitor is required to split off the AC signal. Together with the input impedance of the subsequent stage, this capacitor forms a high-pass. In many circuits, its cutoff frequency is so low that it does not seem to have an effect. For example, in the Fender Bassman (held in highest regard also by guitarists), we find f_g = 3Hz (50nF/1MΩ) which is way below any normal frequency found in the guitar. However, Fig. 10.1.18 shows that this coupling cap has an effect despite its low cutoff frequency: the non-linearity will generate extremely low frequencies (0 Hz if you wait long enough …) that are split off by the high-pass. Taken by themselves, these low frequencies would be inaudible. However, they do determine the position of the operating point and therefore influence the distortion of the subsequent stage. The specific value of the cutoff frequency also has a significance because it determines how fast the transient processes run (Chapter 10.1.6). This example very clearly shows that design rules valid for linear operation can lose their relevance in an overdrive scenario.

In our second example (**Fig. 10.1.19**), a signal of two sine signals close in frequency and beating against each other is first distorted with a tube characteristic, and then subject to hard clipping. The scaling of the ordinate is chosen for all pictures such that equal amplitudes result for small-signal operation. Given such an extreme clipping one could surmise that the "soft" tube distortion occurring first would not have any actual effect since subsequently we have clipping, anyway. As long as there is no coupling capacitor between the two distortion stages, this assumption is indeed correct. However, as a coupling capacitor is introduced, the signal changes – in particular in the low-frequency region and in the area of the summation frequency of the two sinuses (in this example around the 31st harmonic).



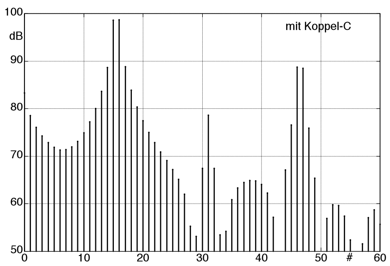


Fig. 10.1.19: Beat-signal (1st column).
Distorted with tube characteristic (2nd column),
then subject to hard clipping (3rd column).
The spectrum given in the 3rd line shows
the effect of the coupling capacitor.

To *round* off this section, let us bring the "round" tube distortions face to face with the typical OP-clipping. If we trust literature, then the latter is the reason for the "harsh" transistor-sound – as opposed to the soft tube sound. Sure, there are differences in the spectrum (**Fig. 10.1.20**), but in fact we also find similarities. In any case, the visual impression ("a round signal shape will sound more round, as well") should not be overrated; tube- and transistor-amps differ in much more than just the rounding of the signal-shapes. Only the connection of several systems makes for the amp. Or, rather, for the sound …

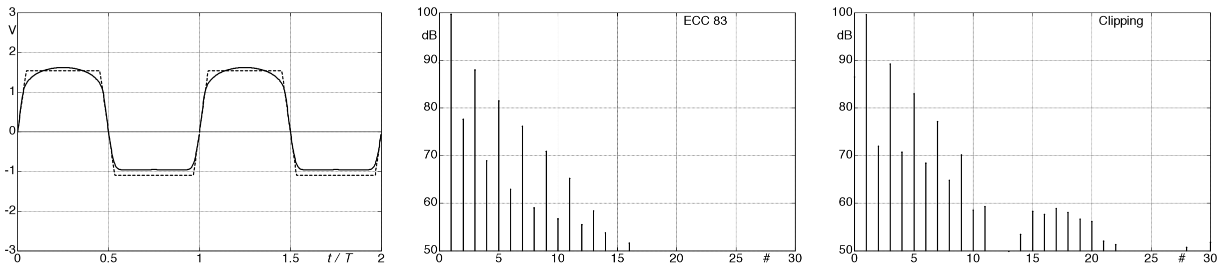


Fig. 10.1.20: Tube distortion (ECC83) compared to hard OP-clipping, driven by a sinusoidal signal; the distortion levels below 60 dB correspond to a harmonic distortion of $< 1\%$ in this example.

Special consideration needs to be given to slew-rate-limiting since such non-linearity does not occur in tube amplifiers. The **slew-rate** *SR* is the speed of change in the signal i.e. the derivative *dU*/*dt*, usually given in V/µs. For a sinusoidal signal, the maximum slew-rate is present at the zero-crossing: $SR = 2\pi f \cdot \hat{u}$. A voltage amplitude of 13V (a typical OP amplitude) results in a slew-rate of just about 1 V/µs at f = 12 kHz. If the maximum slew-rate that the amplifier can provide is smaller than the signal slew-rate, then non-linear distortion results. In contrast to a low-pass (the *linear* transformation of which can alternatively be specified by a cutoff-frequency and a time-constant), the slew-rate-limiting is a *non-linear* transformation that changes the signal shape in particular close to the zero-crossing (**Fig. 10.1.21**).

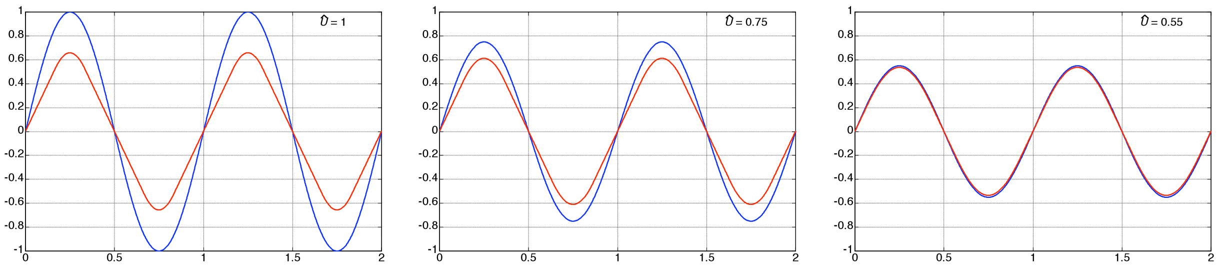


Fig. 10.1.21: Sine-functions of different amplitude (——), non-linear transformation w/slew-rate-limiting (——).

Although in principle the slew-rate may be limited for rising signals to a different value compared to falling signals, both values are almost equal for most operational amplifiers: for example for the (outdated) LM-741: SR_{max} = 0,5 V/µs, or for the TL-071: SR_{max} = 13 V/µs. With a SR_{max} = 0,5 V/µs, the maximum frequency for distortion-free, full-drive-level operation is only 6 kHz. One could assume that this would suffice for a guitar amp since most magnetic pickups limit their spectrum at the most at this frequency. However, this assumption overlooks the possibility of overdrive: if this 6-kHz-tone overdrives the OP by a factor of 10, then the signal-slew-rate is also 10 times as quick at 5 V/µs. **Fig. 10.1.22** shows that slew-rate limiting and clipping are two different kinds of non-linearity: clipping limits the too-large values of the signal while slew-rate limiting confines the value of the slope of the signal. If both types of distortion happen in one and the same stage, the sequence needs to be considered: the two transformations are not commutative!

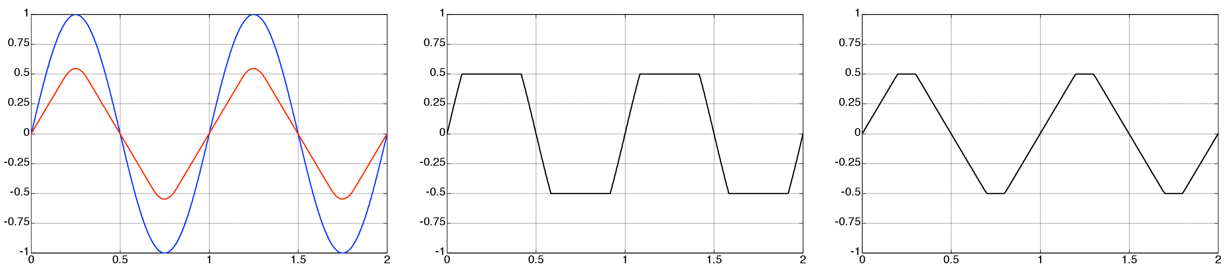


Fig. 10.1.22: Sinusoidal signal (——), slew-rate limiting (—— left). Sinusoidal signal with clipping (middle). Sinusoidal signal with slew-rate limiting and subsequent clipping (right).

The principles of circuit design are the reason that we get slew-rate limiting with an OP but not with a tube (in a comparable manner, anyway). In the OP a number of subsequent stages generate a very high amplification (e.g. 100.000) that is then reigned in by negative feedback to e.g. 50. The same gain is accomplished in tube amps in a single stage without or with very little negative feedback. The high gain of the OP forces another difference: in order for the overall feedback-loop to remain stable, a low-pass characteristic (e.g. with a cutoff at about 100 Hz) is required in the forward branch (i.e. in the pure OP without the feedback network).

Working in the framework of a model, we may replace the typical OP by the building blocks shown in **Fig. 10.1.23**: a comparator (subtractor) is followed by a first amplifier, a first limiter, a 1st-order low-pass (with e.g. the aforementioned 100 Hz cutoff), a second amplifier, a second limiter. A (negative) feedback branch connects the output to the other input of the subtractor). The DC-gain is e.g. 100.000; the gain drops off with $1/f$ from 100 Hz, reaching the value of 1 at 10 MHz (the so-called transit frequency).

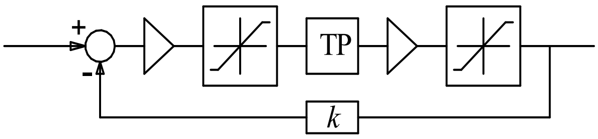


Fig. 10.1.23: Block-diagram of a typical operational amplifier with negative feedback.

Limiting can occur at *two* places in this amplifier with different effects on the external world: limiting in the output stage will introduce clipping, while limiting occurring ahead of the low-pass will introduce slew-rate distortion. We may see the low-pass approximately as an integrator – this will give us an easily understandable model for the limiting of signal rise-times. If the amplifier is driven with a low-frequency signal, the output stage will limit first and we get clipping. With a high-frequency signal, the stage ahead of the low-pass will limit first and slew-rate-type limiting happens.

It really gets interesting for a mixture of tones, e.g. with a two-tone signal consisting of a 1st and a 2nd harmonic (**Fig. 10.1.24**). This signal has the same peak value both on the negative and the positive side and would be symmetrically limited given a point-symmetric limiter-characteristic. However, since the zero-crossings have slopes of different steepness, the slew-rate distortion has a different effect on the two half-waves, resulting in a shifting of the signal: it moves away from the zero-line and becomes asymmetric. In the OP, the negative feedback would immediately become active (the loop gain is indeed very high at low frequencies), and a counteractive offset voltage would result, with the slew-rate limited signal losing its quality of being DC-free, and experiencing a shift towards the negative (middle picture). Now the clipping is added that predominantly limits the negative half-wave – despite the fact that the original two-tone signal is in fact symmetric with regard to the horizontal axis. The processing of the second signal – a superposition of 1st and 3rd harmonic (right-hand picture) – is just as interesting: the slew-rate distortion does not only reduce the signal but distorts it in a non-linear fashion. Still, dents remain at the location of the extreme signal values – in contrast to the effect of pure clipping. These examples show that the slew-rate distortion occurring in OP-circuits has a very different effect compared to pure clipping process that is often seen as the sole reason for distortion. In the typical operational amplifier, slew-rate distortion does, however, not appear by itself but always in combination with clipping.

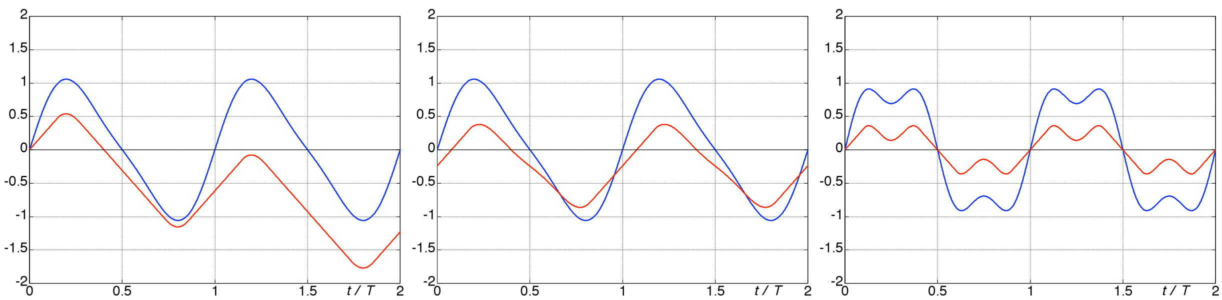


Fig. 10.1.24: Slew-rate limiting: left and middle: 1st and 2nd harmonic. Right: 1st and 3rd harmonic. Two-tone signal (——), slew-rate-limited signal (——).

The maximum slew-rate that an OP can handle may vary dramatically – depending on the OP-type: in this respect the old **µA709** (one of the first universally usable operational amplifiers) was particularly inept. Its maximum slew-rate was (at 0,25 V/µs) so small that full drive was only possible up to 3 kHz at most. Since at the time of the introduction of this OP (1966), harmonic-distortion measurements were usually carried out only at 1 kHz, the slew-rate distortion often remained undiscovered. The **µA741** introduced two years later was able to deal with double the slew-rate but that was still not enough: 10-fold overdrive with a 3-kHz-signal requires 2,5 V/µs. Only later OP-amps – such as the **TL071** at 13 V/µs – reach faster regions. By the way: what is in fact the typical slew-rate of the voltage generated by a magnetic pickup? Of course, this depends on many parameters; in Fig. 10.1.5, for example, 0,06 V/µs is reached. Feeding this signal to a **Music-Man**♣ guitar amplifier, it will be amplified 20-fold in the first OP. To avoid any slew-rate distortion, the OP would have to be able to take on 1,2 V/µs. However, the LM1458 used in some Music-Man amps cannot go beyond 0,5 V/µs without distortion (just like the LM307H used as an alternative). Not all Music-Man amplifiers used these slow LM1458 or LM307H in their input-circuits: some work with the fast TL071 (13 V/µs) … but then feed the signal to a LM1458 in the third amplification stage. Worse: for the input-OP, the gain in the treble range can be increased from 20 to 120 via the "Bright"-switch, increasing the necessary slew-rate value by another factor of six. The distortion generated by this is therefore tube-**un**typical. That the Music-Man amp has a tube power amp ahead of the loudspeaker will therefore not guarantee the same sound compared to an amplifier working exclusively with tubes in its signal path.

Of course, tubes are not infinitely fast, either; however in most cases in tube circuits the rise-time is already limited in the grid-circuit via a low-pass. While this low-pass is non-linear due to its (Miller-) capacitance depending on the voltage-gain of the tube, this non-linearity has an entirely different effect compared to slew-rate limiting.

The following table lists the slew-rate values for some operational amplifiers. Depending on the manufacturer, the numbers differ somewhat: for the LF356, for example, we find both 10 V/µs and 13 V/µs. The first letters in the designation may indicate the manufacturer (e.g. LM 741, or SG 741, or µA 741), while the last letters specify housing types, or temperature ranges, or amplifiers with selected data (e.g. LM 307 and LM 307H). These supplementary letters are, however, not standardized but specific to the respective manufacturer. For some types, the open-loop gain (and thus the slew-rate) can be changed via an externally connected capacitor (so-called compensation, e.g. in the LM 301A).

35 V/µs: HA 5147, OPA 404,
13 V/µs: TL 071, LF 351, LF 353, LF 356,
10 V/µs: LM 302, LM 301A (uncompensated),
6 V/µs: NE 5534, LF 355,
0,5 V/µs: LM 107, LM 207, LM 307, LM 741, µA 748, RC 1458, RM 1558,
0,2 V/µs: OP 07,
0,1 V/µs: LM 108, LM 208, LM 308 (each compensated),

Table: Slew-rates of some selected operational amplifiers (guide values).

♣ Amplifiers and instruments, founded in 1972 by Leo Fender (and Tom Walker), sold to Ernie Ball in 1980.

10.1.5 Frequency limits

The frequency limits of the spectrum of an electric guitar are located at about 82 Hz and about 5 kHz; a guitar amplifier does not have to reproduce any lower or higher frequencies. That is a common and not entirely wrong opinion. The open low E-string vibrates with a fundamental frequency of 82.4 Hz, and the spectrum is limited towards the higher frequencies by the pickup-resonance often located between 2 and 5 kHz. However: bandwidth of the electric guitar and bandwidth of the amplifier are two different things. It is not necessary in a first modeling step to look into the issue that a guitar generates time-limited sounds and therefore the associated spectrum cannot even become zero below 82 Hz. What does require in-depth consideration is the fact that an amplifier with a non-linear distortion-characteristic will generate difference-tones of a frequency far below 82 Hz. Using operational amplifiers (OPs) it would be possible to DC-couple the output of each amplifier stage with the input of the following stage and thus transmit any desired low frequency (down to 0 Hz if we wait long enough …). Such an arrangement is sometimes seen as ideal for recording studio technology because there will be neither phase- nor amplitude errors in the low-frequency region. However, as already previously mentioned: the guitar amplifier is a part of the instrument, it *is supposed* to generate lots of errors. “Errors” from the point of view of classic circuit design, that is, which in the present context are better termed with “signal alterations”. The latter should be of the right kind, i.e. those that sound good – and only those. What sounds good or bad is of course a matter of subjective judgment. If a guitarist wants to hear low-frequency difference-tones, amp and speaker need to reproduce these. This feature is, however, not the norm, because the resulting sound will be assessed by many players as “undifferentiated” and “mushy”. In your typical guitar rig, we therefore see even whole bunch of high-pass filters taking care of effectively attenuating the very low frequencies: several RC-high-passes, the output transformer, and the loudspeaker. An extreme case was already mentioned in Chapter 10.1.4: in some Fender amps, you will find an RC-cutoff as low as 3 Hz. But then there is the other extreme: the 600-Hz-high-pass in the VOX AC-30.

Low frequencies may be attenuated not only in the plate-circuit where the RC-coupling works as a high-pass, but also in the cathode circuit. To obtain the highest possible gain, the cathode-resistor is often bridged by a capacitor. This **cathode-capacitor** will, however, only have an effect as long as its impedance is not significantly higher than the value of the resistor it bridges. Since it is not possible to make this capacitor indefinitely large, two cutoff-frequencies appear: below the lower cutoff-frequency, the capacitor is almost without any effect and the gain here is v_T, while above the upper cutoff-frequency, the gain is v_H, with a monotonous increase in between (**Fig. 10.1.25**).

For the small-signal model, the tube is replaced by an AC-voltage-source of the voltage $U_0 = \mu \cdot U_{gk}$. Here, μ is the open-loop gain of the tube – a theoretical parameter amounting to about 100 for the ECC83. The internal impedance R_i of the tube is connected in series to this source (internally within the tube); for the ECC83, its value is about 50 – 100 kΩ. If we postulate that there is no current though the grid, the plate-current equals the cathode current and is calculated as $I_k = U_0 / (R_k + R_i + R_a)$. I_k generates a negative feedback voltage across the cathode-resistor. The input voltage U_e decreases by the amount of this feedback voltage $U_{gk} = U_e - I_k \cdot R_k$. This enables us to calculate the plate-voltage $U_a = R_a \cdot I_k$:

$$U_a = U_e \cdot \frac{-\mu}{1 + \left(R_i + R_k \cdot (1+\mu)\right) / R_a} = U_e \cdot v_U \qquad \text{Voltage gain (without load)}$$

$\mu = 100$, $R_i = 72$ kΩ, $R_a = 100$ kΩ, $R_k = 1{,}5$ kΩ, yields $v_U = v_T = -30{,}9 \mathrel{\hat{=}} 29{,}8\,\text{dB}$.

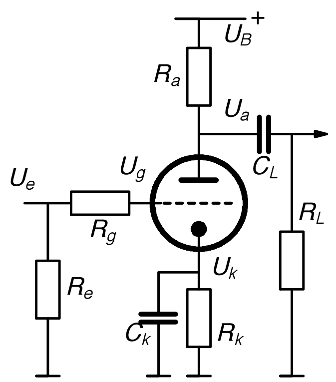
Including the cathode-capacitor (which is taken as a short in the high-frequency region) sets the voltage gain to $v_H = -58{,}1 \mathrel{\hat{=}} 35{,}3\,\mathrm{dB}$; this again is for the unloaded tube. A load resistor is simply connected in parallel to the plate-resistance: with a load of e.g. 100 kΩ, R_a is reduced to 50 kΩ and the voltage gain drops to, $v_T = -18{,}3 \mathrel{\hat{=}} 25{,}2\,\mathrm{dB}$ and $v_H = -41{,}0 \mathrel{\hat{=}} 32{,}3\,\mathrm{dB}$, respectively. For the tube without load, the cathode-capacitor will generate a treble-boost of 5,5 dB, and for the tube loaded with 100 kΩ, the boost will be 7,1 dB (**Fig. 10.1.25**). The capacitance of the cathode-capacitor – in conjunction with the remainder of the circuitry – determines in which frequency-range the transition from v_T to v_H happens. We could surmise that, besides C_k, it is only R_k that sets the treble-boost because this is the resistor that C_k bridges. However, in fact the cathode needs to be considered as load of this two-pole, as well. The relative treble-boost is:

$$v_H / v_T = 1 + R_k \cdot (1+\mu) / (R_a + R_i)$$ Relative treble-boost

The center-frequency f_Z (marked with a small circle in the figure) computes to:

$$f_Z = \sqrt{1 + R_k \cdot (1+\mu)/(R_a + R_i)} \Big/ (2\pi \cdot R_k C_k)$$ Center-frequency

If the cathode-resistor is bridged with a “large electrolytic cap” of e.g. 25 µF or more, the center-frequency is located so low (e.g. 5 Hz) that the gain receives a broadband increase – this being the normal approach for Fender amplifiers. Typical examples for small capacitor values (e.g. 0,68 µF) are found in some Marshall amps (f_Z = 150Hz, ΔG = 8dB).



R_a = 100 kΩ
R_i = 72 kΩ
R_k = 1,5 kΩ
C_k = 0,68 µF
$C_L = \infty$
R_L = 100 kΩ

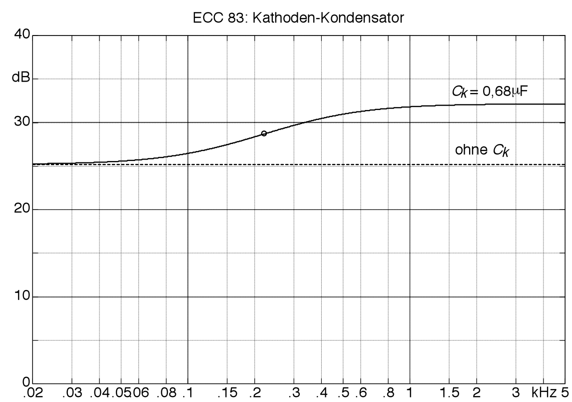


Fig. 10.1.25: Input-circuit of a tube amplifier (left), effect of the cathode-capacitor (right).

In the circuit according to Fig. 10.1.25, the coupling capacitor C_L is taken to be of infinite capacitance in order to be able to show the effect of the cathode-capacitor by itself. In guitar amps, C_L often has a value of 22 nF, but larger values (0,1 µF) are used, as well, as are smaller capacitances (500 pF). When calculating the resulting high-pass cutoff frequencies, it should be considered that the internal impedance of the tube circuit is not zero but is given by the R_a and R_i connected in parallel.

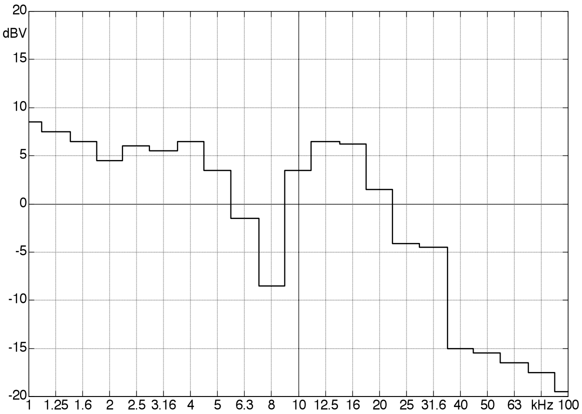
The classic guitar amplifier contains 4 tube-stages and thus has 3 coupling capacitors – the output of power stage is not picked up via a capacitor but via the output transformer. While it is easy to calculate the effect of the coupling caps on the low-frequency-response, the output transformer constitutes a complex system the data of which cannot be seen in the circuit diagram. The upper cutoff-frequency is not apparent, either.

We could assume that the upper cutoff-frequency is always sufficiently high to reproduce the guitar signal (which is low-pass-limited by the pickups), and that therefore it would be not necessary to specially consider it. This assumption is, however, only admissible as long as the amplifier is considered as a linear system. If an overdrive situation occurs, we get signal components in the **ultrasonic range**. These would be inaudible by themselves – however, as ultrasonic signals hit a non-linear amplifier stage, difference tones may be formed that may be audible, after all. A tone-pair constituted of two ultrasound signals (e.g. 24 kHz and 25 kHz) is inaudible at normal levels. Feeding the tone-pair to a 2nd-order distortion-characteristic will generate (among other components) a 1 kHz 2nd-order difference tone that may be audible. This effect should not be seen as all that dramatic, but it should not be entirely disregarded, either. Whether the 1-kHz-tone is in fact audible depends on its level and the levels of further neighboring tones which may have a masking effect. After all, the two ultrasound-tones are not generated in isolation, but are part of a spectrum generated by preceding amplifier stages, and they will not have very large levels. However, since guitar amps may include a very strong emphasis in the high frequency register, a bit of out-of-the-box thinking is advised. We have distortion, treble-boost and subsequently more distortion: there is potential for audible sound differences the reason for which *may* lie in the ultrasonic region.

Why do we not find any **upper cutoff-frequency** in the data-sheets of tubes? Some manuals will give 300 MHz for triodes, or – depending on the type – 1 GHz; however, specifically for the ECC83 this field is usually left empty. The reason is actually rather trivial: the upper cutoff-frequency is determined by the circuitry around the tube. Let's speculate a bit how all this started: the first guitar amps had to be economical regarding the use of power – that made (after the octal-socket-era had passed) the 12AX7 with only 1 mA plate-current highly welcome. As a result, the circuitry had to be of rather high impedance, with 1-MΩ-potentiometers (Fender, Marshall) necessary so as not to load the plate circuits too much. With the center-tap of such a potentiometer set in the middle of the range, its internal impedance is about 250 kΩ[♣]. Connecting this center-tap to the next high-gain triode with an input capacitance (enlarged by the Miller-effect) of about 150 pF, we get a low-pass with a cutoff frequency at about 4,2 kHz. That is kilohertz, not megahertz! You would not want to include such a low upper cutoff frequency in a data-book – it would look quite bad. The relatively high input capacitance is generated by the capacitance between grid and plate (12AX7: C_{ga} = 1,6 pF) that is enlarged by a factor given by the voltage gain. With v_U = 50, this already amounts to 80 pF, and since the wiring leading up to the tube also has a capacitance, 150 pF are easily reached – or even more. The low-pass mentioned above is not always there, though: if the center-tap of the 1-MΩ-pot feeds a cathode-follower (common-plate circuit) the cutoff-frequency will be much higher. In Fenders "Twin-Reverb" (just to name one example), however, the center-tap of the potentiometer directly connects to a common-cathode circuit the input capacitance of which is relatively high. In many Marshall amplifiers there is even a 470-kΩ-resistor in series to the center-tap (summation-stage, total series resistance = 320 kΩ). At this location in the circuit there was also an opportunity to include a low-cost supplement increasing the treble response: a fixed capacitor (Marshall) or a switchable capacitor ("Bright"-switch, Fender). The overall actual cutoff frequency resulting from this hodge-podge of frequency-boosts and frequency-cuts can be calculated via complicated models but depends on many parameters – not least on the layout. The distance between lines leading to grid and plate does influence, via the Miller effect, the input capacity and the upper cutoff-frequency.

♣ The internal impedance of the tube will also make a small contribution.

Fig. 10.1.26 shows a section of the layout of a Fender amplifier (Super-Amp). Resistors and capacitors are soldered to eyelets on a carrier board, and wires lead from the long side of the board to other sub-assemblies (connectors, potentiometers, tubes, transformers). Some wires are laid out below the board at only a small distance to the components above. For example, the wire connecting the grid of a tube is located directly below the coupling capacitor connected to the plate of the same tube – this certainly is not the best possible decoupling approach. Even more extreme is the situation with three wires coming out of an access-hole (in the top section of the picture): two of these are connected to the input jacks, the third carries the plate-AC-voltage of the corresponding input tube. The resulting capacitive coupling is not particularly strong but we need to consider that the grid-plate-circuit is especially sensitive, and that such coupling has the effect of a low-pass. It cannot be excluded that such a low-pass is in fact intended, but comparisons with many other Fender layouts do not really support this assumption. The various wires seem to too arbitrarily keep or change their positions over the years.



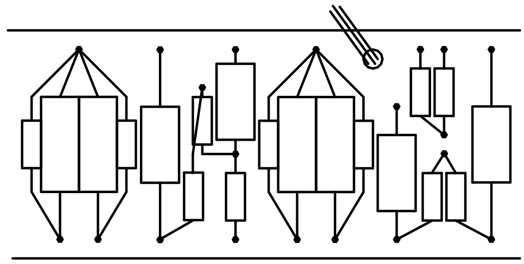


Fig. 10.1.26: Fender component board (excerpt, above).
Third-octave spectrum of the power tube-g_1-voltage (Fender Deluxe). Stratocaster, Stratocaster (left).

As distortion occurs, frequencies above 5 kHz result. The above third-octave-diagram shows this – it is taken from the grid of a power-tube; similar situations can be present at other tubes, as well. The effect of the input capacitance of a tube is shown in **Fig. 10.1.27** using the example of volume-pot: as it is turned down we obtain a low-pass-effect. The cutoff-frequency is lowest at an attenuation of about 6 dB. A “Bright”-capacitor bridging the pot (from the anode to the grid of the subsequent tube) compensates this treble loss but as the control is turned down further, an over-emphasis of the treble occurs. The individual characteristics are strongly dependent on stray-capacitances and on the gain of the individual tubes.

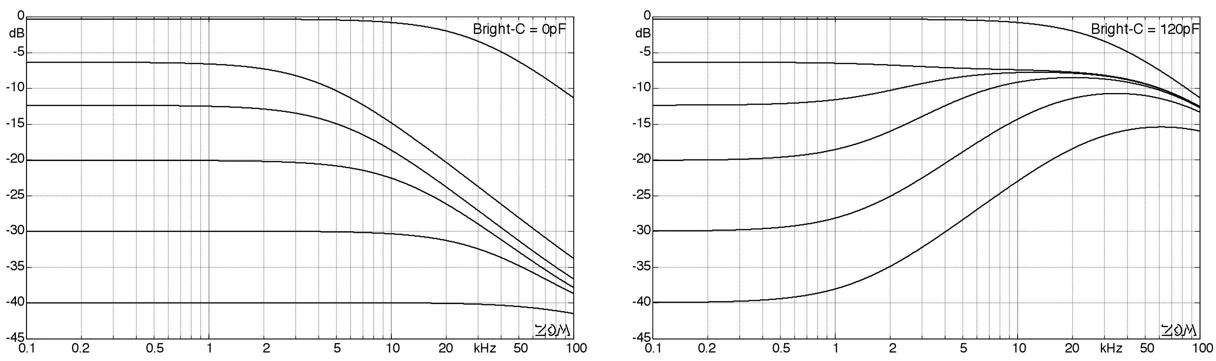


Fig. 10.1.27: Transfer-function of a volume-pot loaded by a capacitance (1 MΩ); tube-input-capacitance 150 pF.

 Translated by Tilmann Zwicker

10.1.6 Time variance

Many theorems of systems-theory are valid only for linear and time-invariant systems. Guitars amplifiers are neither the former nor the latter. There is non-linear distortion, and the characteristics are time-variant: they change due to aging of the components (this being a well-known aspect), and they are subject to short-term shifts of the operating points (this aspect being not recognized much). Trivial time-variances relate to components that change their characteristics dependent on temperature, or that age relatively quickly (such as tubes). These time-variances will not be addressed here – for the present consideration, all components are assumed to be time-invariant. Short-term shifts in the operating point, however, can nevertheless occur, because non-linear processes lead to a re-charging in capacitors. If there were no capacitors, this chapter would be omitted – or the other way round: every capacitor is a potential source for variances.

Tubes in guitar amplifiers are often overdriven – they are non-linear systems. Even for seemingly undistorted (“clean”) sounds, the attack may easily be slightly distorted♣. All even-order distortions (k_2, k_4, k_6 ...) generate an additional **DC-component** that shifts the **operating point** for a short time – the transfer behavior thus becomes time-variant. For example, the **cathode-resistor** is often bridged by a capacitor in order to reduce the negative feedback. The DC-component generated by even-numbered distortion of the cathode current changes – in a time-variant manner – the cathode-voltage and correspondingly the operating point. A further variable is the **supply-voltage** fluctuating (“sagging”) dependent on the power fed to the loudspeaker. While these shifts are low-pass-filtered, they are not regulated out; they have a backwards effect on the plate-voltages of preamp and intermediate amp. In **Fig. 10.1.28**, we see measurements of the supply-voltage of a Fender amplifier (Deluxe). The amp is fully driven from t = 0,3 to 1,7 s and the supply-voltage drops from 247 V to 210 V. As a consequence, maximum signal level and harmonic distortion change as already shown in Fig. 10.1.12. Many guitar players demand this effect since they feel that it renders the guitar sound livelier. However, measures are also taken to reduce this sagging – via changes in the filter capacitors and associated resistors. In early amps, the filter capacitor (C_B) was rather small (8 µF) and was later increased to up to 50 µF. For full removal of the effect, a stabilizer-tube is required. The sagging is not primarily caused by the larger current consumption of the preamp-tube but by the current used up by the power stage that reduces the voltage of the power supply. The exact shape of U_B over time therefore depends on many parameters.

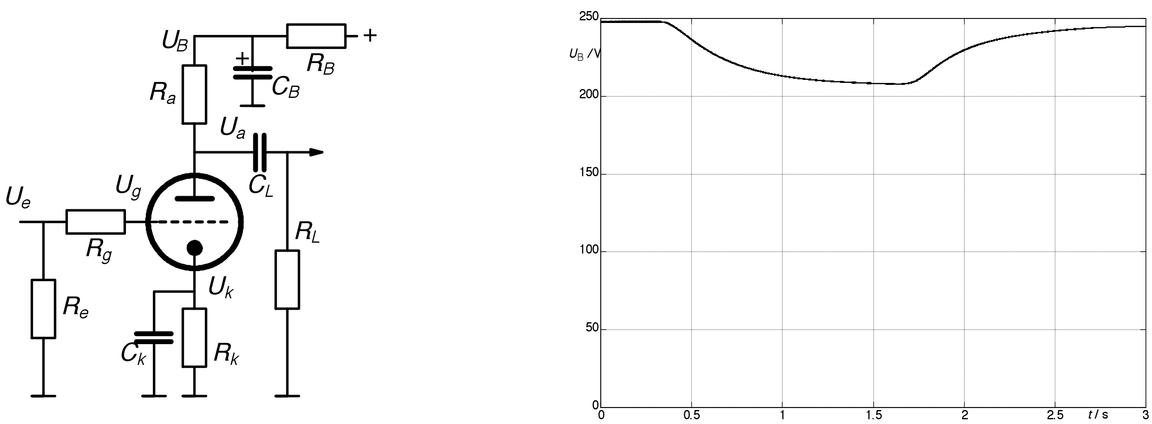


Fig. 10.1.28: Tube input stage (left); drive-dependent sagging of the voltage supply (right).

♣ It is the high art of amplifier-design to make such distortion sound good.

Another variance shows for the **coupling capacitors**. Only for a negative grid-offset does driving a tube not require any power (i.e. no current at the input is required). Typically, we find an offset of $U_{gk} = -1{,}2\text{V}$ in preamplifier tubes, and thus strong pickups can drive the tube into the range of non-negligible grid-current. The subsequent tubes are all the more likely to experience grid-current. Most amplifiers do not use a coupling capacitor at the input; the pickup is usually connected via a resistance of 34 kΩ to the grid of the first tube. If the pickup were connected via a capacitor, only the grid-current could charge it (as a unipolar current creating a negative voltage at the grid) and shift the operating point towards a smaller plate-current. Of course, this is only temporary because the capacitor could discharge via the grid-resistor and the internal guitar-impedance – but exactly these transient processes are NOT present in the input stage of the classic guitar amps (Fender, VOX, Marshall), if we disregard some very early variants using the splash current (i.e. using grid-current bias).

We find a much different situation in the coupling circuits between the individual tubes: here there is almost always a coupling capacitor (the only exception actually being the galvanic coupling in cathode-follower circuits). Since the AC-plate-voltages can be much larger than the voltages allowable for operation without grid-current, a temporary re-charging of the coupling capacitors is almost inevitable. The grid-currents themselves will not in principle lead to audible effects, but the shift in the operating point can lead to audible changes in the harmonic content. We can roughly estimate the speed with which the re-charging processes run: for the "charging" (grid-current flowing) there is a non-linear process because the input-impedance of the tube becomes non-linear. As an approximation, the load-resistance of the preceding tube may serve – in conjunction with the capacitance of the coupling capacitor. Depending on the specific amplifier, the re-charging will happen over the course of a few milliseconds. The "discharging" cannot happen via the grid but only via the leak-resistance (in the order of at least 1 MΩ). This leads to an effect occurring over a time of 20 ms i.e. a time comparable to what is used in studio-compressors (in a "fast" setting). Thus: even if the value of the coupling capacitor is large enough that the high-pass it constitutes is effective only at frequencies far below those of usual guitar signals – the recharging times are defined by these capacitances (and the resistors in the circuit).

Given the sheer variety of tube amplifiers available on the market, it is difficult to specify the typical cause for the "tube sound". Even when only asking about the typical characteristics, a range of different answers is offered; this will happen even more if we look for the corresponding causes. An often heard verdict would be: *the tube amp is alive, it plays more dynamically, sounds more lively, reacts better to changes in the expression*. The opinion regarding a transistor amplifier often is the opposite, it is said to sound *sterile, impersonal, analytical, dead*. The perceived "liveliness" connected to the tube may well have its base in the shifts of the operating point as described above. Even if a unidirectional current as small as 10 µA flows through a 22-nF-capacitor for a mere 1 ms, the resulting voltage change will be 0,45 V. Such a shift in the operating point would drastically alter the transmission behavior of an ECC83. It is not that such a behavior would not be possible to obtain with a semiconductor amplifier, as well – however "modern" circuit design sees big advantages in direct coupling between amplification stages (i.e. without capacitors). That is indeed a conducive approach to minimize artifacts, but in guitar amplifiers that is exactly NOT the issue (or at least not a main one).

P.S.: The term *time-variant* chosen here is valid for short-term considerations; in the long term the shifts in the operating point as discussed above are indeed time-invariant i.e. they run an identical course given identical excitation. This distinction is, however, only important in a strictly systems-theory-oriented approach.

10.1.7 Noise, hum, microphonics

The noise of the input-amplifier tube can be modeled with good approximation by a noise-source connected in series with the tube grid and generating **white noise** with a noise-voltage density of about 5 $\text{nV}/\sqrt{\text{Hz}}$ (compare to Chapter 5.12). Note that this model does include only the stochastic component of the overall interference, and not the hum component generated by the AC-heater of the tube. Moreover, in a typical input-circuit it is not only the tube itself creating the noise: in addition, there is the grid-resistor (34 kΩ) resulting from the parallel connection of the two 68-kΩ-resistors in the input-circuit. In fact, this resistor is the actual culprit and acts as the main noise-source with a model voltage-density of no less than 24 $\text{nV}/\sqrt{\text{Hz}}$! Consequently, it is pointless to consider tubes with lower noise as long as we cling to the classical input-circuit. By the way: the overall noise-voltage may not be calculated by simple summation because the signals from noise-sources are not correlated. Rather, a square-root summation needs to be performed:

$$U_\Sigma = \sqrt{U_1^2 + U_2^2}; \qquad e_\Sigma = \sqrt{5^2 + 24^2}\ \text{nV}/\sqrt{\text{Hz}} = 24{,}5\ \text{nV}/\sqrt{\text{Hz}}$$

Clearly, the noise from the tube contributes almost nothing to the overall noise. However, before taking out the **grid-resistor** and connecting the pickup directly to the grid of the input tube, you should consider that this resistor does have some other jobs to do, too: it limits the grid-current and influences the non-linear distortion of the preamp-tube. Moreover, together with the input capacitance, it does form a low-pass that suppresses unwanted RF (*This is Radio Free Europe ...*). In many cases the noise generated by the grid-resistor will be less than the noise generated by the guitar circuit; the latter may certainly reach voltage-densities of 40 $\text{nV}/\sqrt{\text{Hz}}$ (or even more) in the frequency-range important for the hearing system.

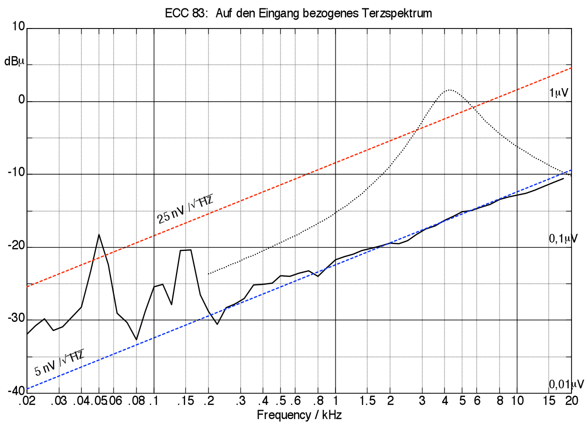


Fig. 10.1.29: 1/3rd-octave noise-spectrum (ECC83). The two dashed lines mark the spectrum belonging to white noise; the dotted line shows the typical noise-spectrum generated by a Stratocaster. All spectra are referenced to the tube input.

In **Fig. 10.1.29** we see the measured third-octave spectrum of an ECC83 in comparison to the theoretical characteristics. For the measurement, the grid was shorted to ground and the tube received DC-heating. Hum of around 0,1 µV is typical for simple shielding; this is much less than the interference caught by magnetic pickups. Without the grid-resistor, the tube creates – across the whole frequency-range – less noise than the pickup measured for comparison (Chapter 5.12). Including the grid-resistor, the pickup noise dominates only in the range of the pickup resonance. The third-octave levels measured at the plate are, compared to the levels given in the figure, larger by the gain factor (33,4 dB in our example). The **broadband** input-noise voltage below 20 kHz amounts to about 1 μV_{eff} (with shorted grid); this is equivalent to a noise-voltage of about 47 μV_{eff} at the plate.

Every guitarist can find out for him/herself which noise-source dominates in a given guitar-amp setting: just compare the noise with shorted input to the noise that occurs with the guitar plugged in and fully turned up. In case both signals are approximately equal in strength, one needs to indeed question the quality of the input tube (or that of the amplifier concept); if there is more noise with the guitar turned up, the interference is caused there. How can we achieve a **short circuit at the input**? The best way is to use a plug with both connecting pins soldered together. Alternatively, a metal potentiometer shaft (6,3 mm) or a similar short-circuit-pin could be plugged into the input jack. Or, very simple: plug in the guitar and turn the volume control (on the guitar) to “0”. Note, however, that this works only if the guitar cable actually reaches the center-tap (middle) connector of the pot (as is the case for Strats or Les Pauls with the customary circuitry). Instruments that have the pots in the so-called “reverse” connection (such as the Fender Jazz Bass) are not suitable for this approach

The second unwanted signal generated in an amplifier is **hum**. It is caused by the power system (230V/50 Hz, or 110V/60, or other voltages/frequencies depending on the country) that contaminates the more sensitive circuit sections via capacitive or inductive coupling[♣]. Faulty design of the layout of the ground-connection can be a reason, as well – especially in the power-rectifier circuit. In the typical tube amplifier we have relatively strong heating currents (preamps tubes: 0,3 A, power tubes 1 – 2 A) the magnetic fields of which can feed into the sensitive plate circuits. DC-heating would be an option for (the customary) indirectly heated tubes but is implemented rarely. It is not really necessary, either: using twisted wiring for the heating and a correct layout of the (electric) ground, every tube amplifier can be constructed in a sufficiently hum-free way such that – for normal use – the hum caught by the magnetic pickups of the guitar dominates.

Microphonics is a term characterizing the tendency of a tube to react to sound (i.e. mechanical vibrations), whether transmitted via air, or structure-borne. Combo-type amps – with loudspeaker and amplifier housed in the same cabinet – are particularly prone to associated problems. The amp may sound as if there is always a bell operating in the background, and at high volumes a howling, uncontrollable feedback may occur. The cause of microphonics is a deformation in the tube-interior, in particular in the (control) grid. The ultra-thin grid wires start to vibrate as sound impacts on the tube, and this in turn modulates the plate-current and generates interfering noises. Every tube is microphonic – but not always to the extent that problems result. Preamp tubes with their very small signal voltages should have especially low microphonics, and tubes specially selected towards this goal are available.

In an orientating **measurement**, a double-triode (12AX7) that generated a clearly ringing tone at 630 Hz when tapped was subjected to sound coming from a loudspeaker. At 130 dB SPL (a sound pressure level easily reached in a combo), an interference voltage of about 1 mV (when referenced back to the input) occurred. A 12AU7 was even considerably worse at 30 mV! Even without fully turning up an amp, such a tube will start to bring some undesirable accompaniment, and feedback whenever the amplification is high. Vibration can get to the tubes not only via air but also via the tube-socket. Consequently it is advisable to consider – at least for the preamp – mounting the respective tubes in sockets using rubber or a similar mechanical absorbent material. The latter should be able to withstand heat while not being prone to embrittlement.

[♣] This happens not only at 50/60 Hz but also at the multiple frequencies, i.e. at 100/120 Hz, 150/180 Hz, etc.

10.1.8 Noise processes

Noise belongs to the stochastic signals – it is not possible to predict its exact course. The simplest quantitative description involves RMS-value, bandwidth, and spectral-envelope characteristic (dP/df = const. or $1/f$). Supplementary specification regarding the time function is given by distribution (= probability-density-function) and cumulation (= probability-density-distribution), further information regarding the spectral distribution results from DFT- and 1/3rd-octave-sprectra. The literature listed at the end of this chapter may serve as guide to the theoretical principles for the description of random signals – the following listing in short introduces the most important noise processes.

a) Thermal noise (white noise i.e. dP/df = const.)

The temperature-dependent random-movements of free charge-carriers in a conductor (or in a resistor) lead to a thermal open-loop voltage at the connecting terminals (without any load); the RMS-value of this voltage is computed to:

$$\boxed{\widetilde{U}_n = \sqrt{4kT \cdot \Delta f \cdot R}, \quad e_n = \sqrt{4kT \cdot R}} \qquad 4kT = 1.70 \cdot 10^{-20}\ \text{Ws}, \quad T = 308\text{K}$$

Open-loop noise-voltage density e_n and RMS-value of open-loop noise-voltage $\widetilde{U}_n$ for Δf = 10 kHz at resistor R:

R =	58.8	100	200	1k	10k	100k	1M	Ω
e_n =	1.00	1.30	1.8	4.1	13.0	41.2	130	$\text{nV}/\sqrt{\text{Hz}}$
$\widetilde{U}_n$ =	0.1	0.13	0.18	0.41	1.3	4.12	13	µV

b) Shot noise (white, i.e. dP/df = const.)

Shot noise occurs in semiconductors and amplifier tubes. It is caused by statistic fluctuations of the current-flow through an interface layer between potentials. As an example, the electron-emission at an amplifier cathode may be modeled by a Poisson-distribution, with the current not continuously flowing but having statistic fluctuations. The real tube-noise is (given the space-charge conditions) slightly less than the theoretical maximum value calculated below for saturation [Meinke/Gundlach]:

$$\boxed{\widetilde{I}_S = \sqrt{2e \cdot \Delta f \cdot I_0}, \quad i_S = \sqrt{2e \cdot I_0}, \quad \widetilde{U}_S = \widetilde{I}_S \cdot R} \qquad 2e = 3.204 \cdot 10^{-19}\ \text{As}$$

Noise-current density i_S, (RMS) noise-voltage $\widetilde{U}_S$ across a 10-kΩ-resistor for 10 kHz bandwidth, generated by DC I_0: [f = Femto = 10^{-15}, p = Pico = 10^{-12}]

I_0 =	10 n	100 n	1 µ	10 µ	100 µ	1 m	10 m	A
i_S =	56,6 f	179 f	566 f	1,79 p	5,66 p	17,9 p	56,6 p	$\text{A}/\sqrt{\text{Hz}}$
$\widetilde{U}_S$ =	56,6 n	179 n	566 n	1,79 µ	5,66 µ	17,9 µ	56,6 µ	V

The relation between shot-noise voltage $\widetilde{U}_S$ and thermal noise-voltage $\widetilde{U}_n$ depends on the DC voltage across the resistor and on the temperature voltage:

$$\boxed{\widetilde{U}_S / \widetilde{U}_n = \sqrt{U_0 / 2U_T}}$$

U_0 is the DC-voltage across resistor R; $2U_T = 2 \cdot 26$ mV = 52 mV.

c) Flicker noise (approximately pink, i.e. $dP/df \sim 1/f$)

This is low-frequency $1/f$-noise caused by inhomogeneities in the material, deficiencies from manufacture, contaminations, and charge-fluctuations at surfaces. The designation stems from the burn spots jumping around (flickering) on the cathode of an amplifier tube. Simplified, the power-density decreases towards high frequencies with $1/f$ (pink noise). However, also observed were noise processes the spectral density of which does not correspond exactly to the $1/f$-hyperbola. Flicker noise is only relevant in the low-frequency range.

The $1/f$-noise caused in **resistors** carrying DC is characterized by the **Noise-Index** *NI*. Metal-film resistors (homogeneous crystal lattice structure) feature a small *NI*, while carbon composition resistors have large *NI*-values. In general, resistors with a high power-handling capacity (and requiring a larger volume) generate less noise than their low-power cousins of the same basic build.

$$\boxed{NI = 20 \cdot \lg \frac{U_{10} / \mu\mathrm{V}}{U_0 / \mathrm{V}} \mathrm{dB}} \qquad U_{10} = U_0 \cdot 10^{-6} \cdot 10^{NI/20\mathrm{dB}}$$

U_0 represents the DC-voltage across the resistor, U_{10} is the resulting $1/f$-noise-voltage (RMS value) per frequency-decade; $NI = 0$ dB $\Rightarrow$ 1 µV/V.

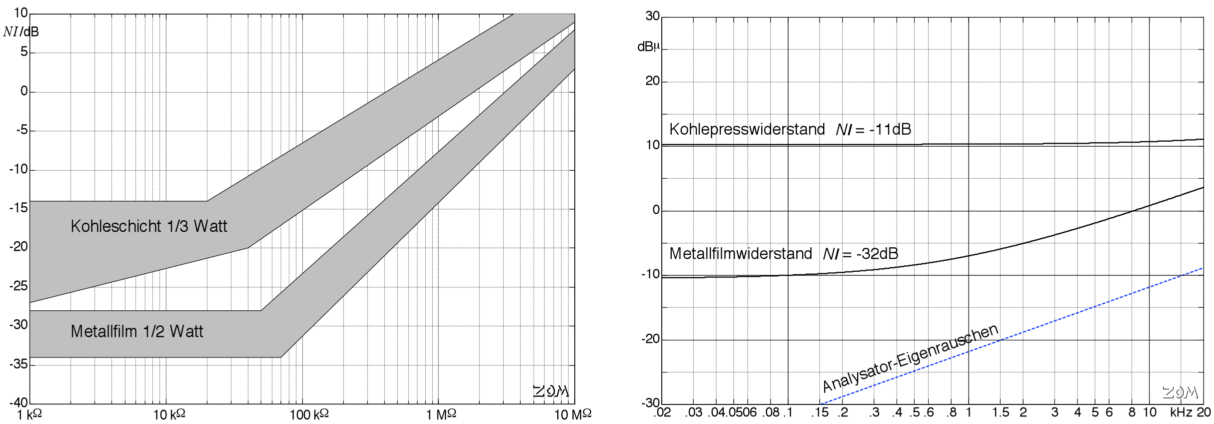


Abb. 10.1.30: left: noise-index *NI* for two different resistor types (Kohleschicht = carbon layer, Metallfilm = metal film). The grey areas show the scatter range between typical average values and typical maximum values. Right: measured 1/3rd-octave noise-; dashed: intrinsic noise of analyzer. Pink noise results in frequency-independent 1/3rd-octave-level voltage levels; for white noise the 1/3rd-octave-levels rise with 10 dB/decade. Kohlepresswiderstand = carbon-composite resistor; Metallfilmwiderstand = metal film resistor; Analysator-Eigenrauschen = intrinsic noise of analyzer.

In **Fig. 10.1.30**, *NI* is listed for different resistor types. The areas marked in grey can only give very approximate orientation-values since the individual build has significant influence on the *NI*. In the right-hand section of the figure, we see measurements taken with two serially connected 68-kΩ-resistors carrying a DC of 1 mA. The two incoherent noise currents of the two resistors need to be added via a Pythagorean summation, and the mutual loading plus the loading via the analyzer (100 kΩ) has to be considered, as well. The metal-film resistors show a thermal white noise in the high-frequency region, and a current-dependent pink noise at low frequencies. In the carbon-composite resistors, current-dependent pink noise dominates throughout practically the whole frequency-range. These measurements give a noise index of the carbon-composite resistors of -11 dB, and an *NI* for metal film resistors of -32 dB. At low frequencies, the noise power densities of these two resistor-types therefore differ by **a factor of 126**. This factor is current-dependent; 1 mA is typical for plate-currents in preamplifiers.

Despite this considerable current-noise, carbon-composite resistors are listed as "**absolute high-end**" in the catalog of a retailer; one is very temped to interpret this as "absolute upper range of the resistor noise". The high-end fan must furthermore not be irritated by the fact the carbon-composite resistors have also considerably larger tolerances (compared to metal film resistors): maximum ±10% (carbon) vs. maximum ±1% (metal). Measurements confirm this: + 7% (carbon) vs. -0,3% (metal). What about the price-difference? As expected, carbon-composite resistors are about 10 times as expensive as metal film resistors. Say no more: it's about more noise – more tolerance to resistance – more money …

What remains is the question whether differences in the current-noise play any role at all compared to the **shot-noise** generated in the tube. For an **ECC83** (12AX7), the equivalent input-noise voltage-density may be set to about 5 $\mathrm{nV}/\sqrt{\mathrm{Hz}}$ as a good approximation. With a voltage gain of 34 dB, this is equal to 250 $\mathrm{nV}/\sqrt{\mathrm{Hz}}$ at the plate, corresponding to a third-octave-level of 11.6 dBµ at 1 kHz (bandwidth 232 Hz). In comparison, the thermal noise from the **grid-resistors** (68 kΩ // 68 kΩ = 34 kΩ) typically found in the input-stages of guitar amps is five times as much (Chapter 10.1.7), reaching some ample **26 dBµ** in the third-octave band around 1 kHz. And how are we doing regarding the resistor-noise created by the plate-current? Given a 100-V-voltage-drop across the plate-resistor, and including a noise-index of NI = -11 dB, we would be confronted with a 1-kHz-third-octave-level (open loop) of no more than 19 dBµ. With the loading by the internal impedance of the tube, this would decrease to about **11 dBµ**. Consequently, the current-noise of a carbon plate-resistor (NI = -11 dB) at 1 kHz is lower than the noise of the preamplifier by 15 dB. For higher frequencies, this difference will grow even bigger, and only below 31 Hz, the current-noise would become dominant for the present model.

So: The current-noise of customary carbon resistors is inaudible in the investigated circuits
But: Supposedly there are carbon-composite resistors with *NI* not at -11 dB, but at 0 dB,
Or even higher – that could then just become audible.
Question: Is that worth 10 times the price? *Answer:* sure, the retailers are happy.

Two advantages are often highlighted to scientifically support the apparent superiority of the carbon-composite resistors: high power capacity with impulses, and small inductance. There may be scenarios in which the relative long thermal time-constant of the carbon-composites helps to avoid overheating, but pre-amp stages in guitar amps are not even remotely in the playing filed here. O.K. then: the reported low **inductance** of composite resistors will be crucial, won't it? No, sorry, that aspect is utterly insignificant in the relevant frequency-range! The impedance of a 100-kΩ-resistor will increase by 0,000000002% at 100 kHz (with a inductance of 1 µH as a baseline). This increase should be seen relative to the manufacturing tolerances in carbon-composites: 10% according to data sheets. Plus: do not forget that 1 µH is already a high value; in data sheets we often find the entry "a few nano-henry". BTW, our metal-o-phobic friends prefer not to mention capacitive reactive values, although these exist in carbon resistors, as well. Do you need to consider those? Course you do … if you want to look beyond 1 MHz, where the reactive currents start to achieve some significance.

The never ending Internet saga of Carbon Comps: "Smooth, creamy sound…Are unstable, should not be used…Very clean and natural sound…Should be avoided…Taut and 3-dimensional sound…Make the working point drift away…Are the only choice for guitar amps…Never heard any difference in sound…Light-years ahead." More examples are available …

Literature: Motchenbacher/Connelly: Low-Noise Electronic System Design, Wiley 1993. **Connor**: Rauschen, Vieweg 1987. **Hänsler**: Statistische Signale, Springer 1991. **Bendat/Piersol**: Random Data, Wiley 1986.

10.1.9 Pentode-preamp

Of all amplifier tubes, the pentode is the most widely used. Compared to the triode, the technical advantage of the pentode used in input stages is based on the high internal impedance and the very small capacitance between grid and plate. Large voltage gain is possible without running the danger of self-excitation [Meinke/Gundlach]. However, as we so often see it: what is true for classical circuit design does not necessarily hold as guideline for guitar amplifiers – the latter most often employ a triode in the first stages. There are exceptions, though: the VOX AC-15 or the Fender Champ may serve as examples. In these rather early amps, we find a **pentode** in the preamplifier. We will look into the technical details of this five-electrode-tube a little later; as a simplification, it functions similar to the triode: the plate-current is controlled by the voltage at the control-grid, the extra screen-grid (g_2) is connected to a constant (high) voltage, and the suppressor-grid is joined with the cathode. The **transconductance** of the **6 SJ 7** pentode used in the Champ is rather comparable to that of an ECC83 (1,6 mA/V) but the internal impedances are very different: 1000 kΩ in the 6 SJ 7 but merely 63 kΩ in the ECC83. Purely by way of calculation, this yields – e.g. for R_a = 200 kΩ – an operational gain of 267 (6 SJ 7) and 48 (ECC83). The operational gain-factors therefore differ by 15 dB! The **EF 86** as it is deployed in the AC-15 features even larger values for transconductance (2 mA/V) and internal impedance (2500 kΩ), and we get an additional 3 dB gain.

It was the susceptibility to oscillations that made VOX replace the EF 86 by a triode, after all: *The EF 86, although excellent electronically, was susceptible to mechanical damage through vibration and would soon begin adding it's own ringing, rattling accompaniment [Petersen/Denney]*. Another reason could lie in the seeming advantage of the pentode: its high voltage gain is helpful when dealing with small input signal. However, when confronted with pickups able to deliver in excess of 1 V, this advantage can easily backfire: the preamp will generate considerable distortion that is not generally desired.

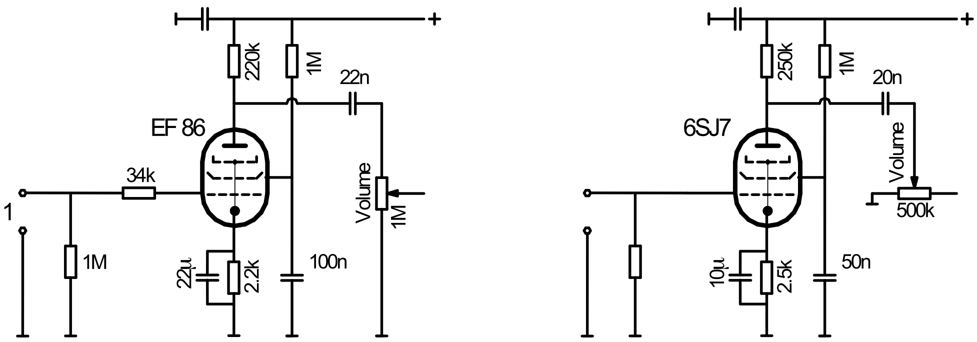


Fig. 10.1.31: Pentode-input-stages in guitar amplifiers: VOX AC-15 (left), Fender Dual Professional (right).

Fig. 10.1.31 shows the input-circuits of two early guitar amps. The AC-15 employs the more modern pentode with the noval-socket while the Dual Professional (developed more than 10 years earlier) still relies on the octal-tube. Only shortly thereafter Fender changes to the dual-triode 6SC7, and in the following generation to the 12AY7.

10.2 Intermediate amplifier

In the signal path, the intermediate amplifier operates between input stage and power stage – but it is not the tube immediately ahead of the power tubes that is meant here (that would be the tube commonly known as *phase inverter* in push-pull arrangements, see Chapter 10.4). In the classical guitar amp, the typical intermediate amplifier is the second amplification stage. Between the first and the second stage we find the tone-filter … or the volume control … or both. In fact even in the classic amp-forefathers we already find different concepts.

Which are the (dis-) advantages of these various topologies, and what are the sonic differences? That is quite difficult to answer. It is easier to address the question what the reasons could have been to implement the respective topology. **Fig.10.2.1** shows the most important ones – there are more but we will not investigate them here. In almost all guitar amps, the signal from the pickup is directly fed to the first tube. This is because any circuitry connected between pickup and the first tube would have to be high-impedance and thus would unduly increase noise. If the volume control is located directly after the first tube and the tone-filter subsequently (as shown in the first variant), then the source circuit (the volume control) feeding the tone-filter would have an internal impedance that depends on the position of the potentiometer's center-tap. Moreover, the potentiometer load (= filter-input) would be frequency-dependent. The effect of the filter would therefore not only depend on the settings of the tone control, but also on the setting of the volume control. Presumable it was this interdependency that precluded the corresponding topology from become really widespread.

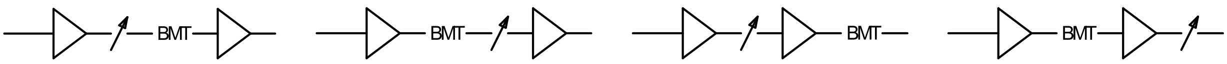


Fig. 10.2.1: Some obvious circuitry-topologies. BMT = tone-filter, arrow = volume-pot.

We will look more closely at the second and third topology-variants; these two are found most often in tube amplifiers. The fourth variant would work without any problem, as well, but was apparently not seen as directly superior and was thus rarely used. Variant two and three differ in the position of the tone-filter: ahead of the volume control or after it. The **sequence** of the subsystems in a guitar amp would be rather unimportant if the amp were a linear system. However, as Chapter 10.1 has shown, non-negligible harmonic distortion happens as early as the very first amplifier stage; the system is non-linear in quite a complicated way. Moreover, a further non-linear effect needs to be considered: the noise that every component generates. Non-linear system need to be source-free i.e. they must not include any noise-sources, either. If the volume-pot is positioned late in the signal-flow (close to the power amp), almost no noise will be audible when the volume control is turned down. However, there is now considerable danger that one of the preceding amplifier stages will be overdriven in case the connected guitar has a high-output – and this danger cannot be reduced by turning down the volume control. If, conversely, the volume control is located directly after the first stage, any potential overdrive of subsequent stages is fully controllable – but there may be a considerable noise level even with the volume set to zero. Of course, no guitarist plays his or her amp with the volume fully turned down so this would probably not be a problem. Rather, the sales department that makes demands here: in the music store, it's no good if the amp creates such a racket even though no-one is even playing through it. Still, the amp needs to be "clean" at low volume. Only later amplifier generations include "Fat"- and "Boost"-switches, and master-volumes to get more sound-options; the early amplifier-variants had to do without that. Obviously, "sound" won out over "noise": in the circuits, the volume control was close to the input (mostly before the second tube stage).

10.2.1 Intermediate amplifier in common cathode-circuit

The standard version of the intermediate amplifier contains *one* tube (almost always a triode) in common-cathode configuration. The circuit is similar or even identical to the first preamp-stage. And why not – the signal has been attenuated by tone-filter and/or volume control and needs to be re-amplified, with the common-cathode configuration being highly suitable. Sometimes, the developers see a need for an impedance conversion in the second amplifier stage – this aspect we will cover in the next section (10.2.2).

In the **common-cathode circuit**, the cathode is connected to "common" i.e. to ground. The required grid-offset is usually generated "automatically" by a cathode-resistor (Chapter 10.1). A capacitor is connected across this resistor in order for the latter to be active only for DC, and to avoid any AC-voltage across it (which would introduce negative feedback). As long as there is not grid-current, this circuit features very high input impedance – although a non-negligible input capacitance (100 pF minimum) does require consideration. The output impedance (internal impedance) results from the parallel connection of the internal impedance of the tube (about 60 kΩ) and the plate-resistor (100 kΩ); the gain factor is about 35 dB (or a bit less if there is significant loading).

Fig. 10.2.2 shows two famous amplifier concepts in comparison: in the Fender circuit, the volume potentiometer directly follows the tone-filter and feeds the intermediate stage, while in the VOX, the intermediate stage is placed between volume pot and tone-filter. **Fender** follows the simple line of thinking: take care of all control efforts at one and the same location. The interaction between the directly connected volume control and tone-filter remains within reasonable limits because the pot is of relatively high impedance (1 MΩ). With the **VOX**, we find an entirely different approach: a special intermediate amplifier with high-impedance input (common cathode configuration) and low-impedance output (common-plate configuration, see 10.2.2) follows the volume pot.

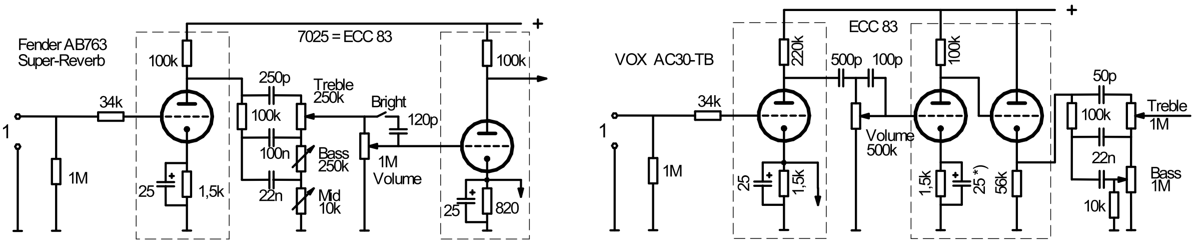


Fig. 10.2.2: Comparison between a typical Fender-circuits (left) and a VOX-circuit (right).
***)** There are VOX amps that do not include the cathode-capacitor for the 2nd tube.

Pushing the discussion of the tone-filter into Chapter 10.3, we will first analyze the 2nd tube-stage of the **Fender circuit**. Both 1st and 2nd tube-stages are fundamentally similar but there are differences regarding the cathode circuit: in the Super Reverb (under scrutiny here), the cathode-RC-circuit also feeds the corresponding cathode of a tube in the other input-channel. Other Fender amplifiers include the same component-saving detail. In the figure, the second tube is not included but an arrow indicates the connection to it. For the grid-offset of the tube(s) to remain at the desired value, the value of the cathode-resistor common to both tubes is approximately halved at 820 Ω (instead of 1,5 kΩ). Since both triodes are feeding relatively high impedance circuits, they have similar voltage gains. Given a regular ECC83, each triode will yield about 32 – 34 dB. The harmonic distortion, however, will be different because the source impedances (ahead of the grid) differ.

10.2.2 Intermediate amplifier with cathode-follower

The VOX-circuit (Fig. 10.2.2) differs from the Fender-circuit not just in the sequence of the partial systems but also in the build of the second amplifier stage. It deploys *two* triodes: the first generates the required voltage gain while the second acts as a current amplifier (impedance converter = cathode-follower = common-plate circuit) and achieves a low output-impedance (= internal impedance). Strictly calculating the internal impedance according to text-books we get $1/S$ (S = transconductance); for the present circuit this would be 600 Ω. An output-impedance of such a low order would not be mandatory, though: the load imposed by the VOX-tone-filter is always larger than 100 kΩ. Before we go into further detail regarding the rather special dimensioning of the VOX-circuit, let us quickly review the history of the cathode-follower: Leo Fender outfits his tweed amps with this circuit from the mid-1950's (albeit not using the 12AX7 but the 12AY7, **Fig. 10.2.3**).

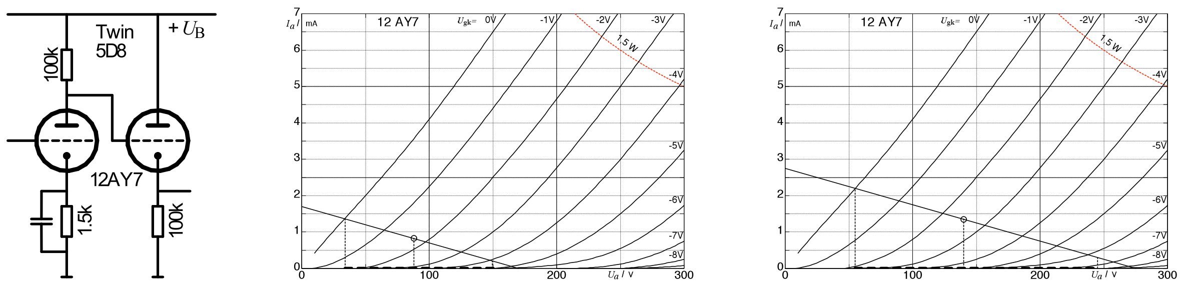


Fig. 10.2.3: Intermediate amplifier with cathode-follower; family of output-characteristics of the 12AY7, U_B = 170V … 275V.

For the 5D8-Twin, the layout specifies [Funk] a supply-voltage of U_B = 170 V, for the later 5E6-Bassman this has risen to 235 V, and in the 5E6-A we find even 275 V. With the increase of the supply-voltage, the quiescent current of the triodes also mounts; this is indicated in Fig. 10.2.3 as a dot on the load-line. For U_B = 170 V, the travel of the plate-voltage of the first triode is limited to about 35 V towards small values by the U_{gk}=0V-characteristic. For even smaller U_a (i.e. larger I_a), the grid would have to become positive relative to the cathode but this is only possible to a small extent: the grid-current is kept low by the high-impedance feed. If the first tube were in blocking mode, its plate-voltage would be the same as the supply-voltage (with no load present). However, since in the second triode there is a grid current (200 μA), the plate-voltage of the first tube rises only to about 150 V. Corresponding characteristics result for a supply-voltage of 275 V (**Fig. 10.2.4**).

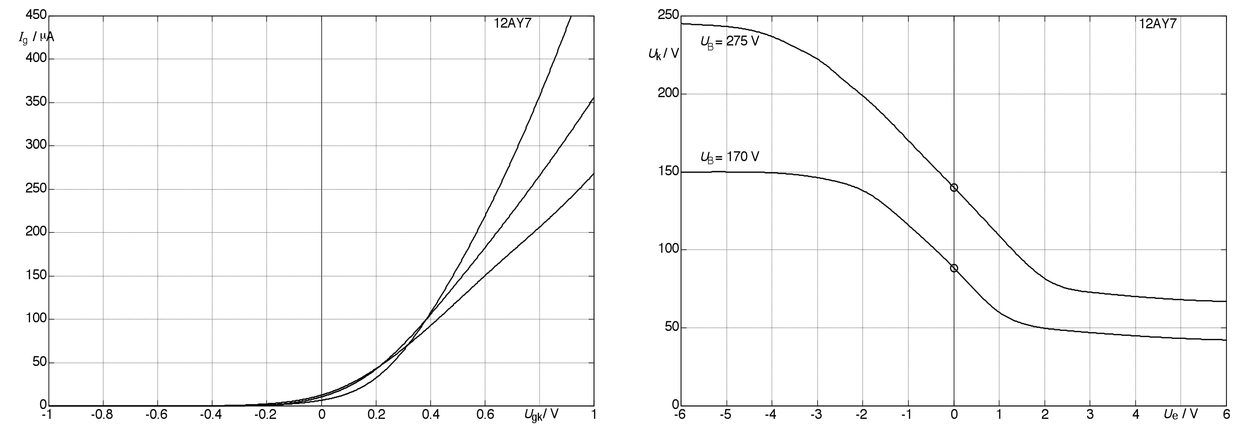


Fig. 10.2.4: Grid-current (left, measured for three different tubes); transmission characteristic (right). The first tube is driven via a 100-kΩ.grid-resistor.

With the change from the E- to the F-series, Fender replaced the **12AY7** by the **12AX7** (= 7025 = ECC83) – presumably because the latter has higher gain, or simply in order to standardize. Bassman 5F6, Super 5F4, and Twin 5F8 still included the common-cathode/plate-circuit for their intermediate amplifiers but received the 12AX7 instead of the 12AY7. In the Super 5F4 the associated components remained identical; for the other amps R_{k1} was decreased from 1.5 kΩ to merely 820 Ω. The differences between the two double triodes are shown in **Fig. 10.2.5**: the 12AX7 sports the larger open-loop-gain (μ = 100 vs. 44) but also has the larger internal impedance: 63 kΩ vs. 25 kΩ. Since the tubes are not operating under open-loop conditions, the gain in reality differs not that much but still considerably: 50 vs. 30, i.e. 34.0 dB vs. 29.5 dB.

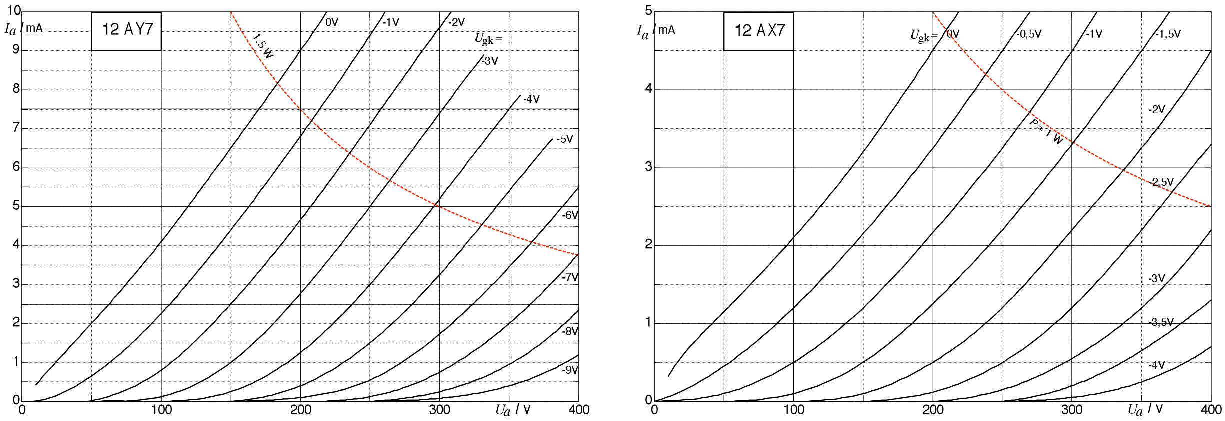


Fig. 10.2.5: Output characteristics (according to data sheets) of the 12AY7 (left) and the 12AX7 (right).

The transmission characteristic of the 5F4-circuit is shown in **Fig. 10.2.6**. Besides the steeper slope (= higher voltage gain) it is especially the much stronger curvature that stands out – it is the reason for strong non-linear distortion. The change to the smaller cathode-resistor (5F6) balances the operating point somewhat but cannot change anything about the curvature. It may be due to this non-linear behavior that Fender's Super-Amp 5F4 received additional negative feedback – but the Bassman 5F6 (and its successor 5F6-A) had to do without the negative feedback. It needs to be noted that in particular this Bassman had a lasting influence on the British amplifier industry: it was the amp that Jim Marshall modeled his JTM amps after from 1962 (with cathode-follower, with 820-Ω-resistor, without additional negative feedback).

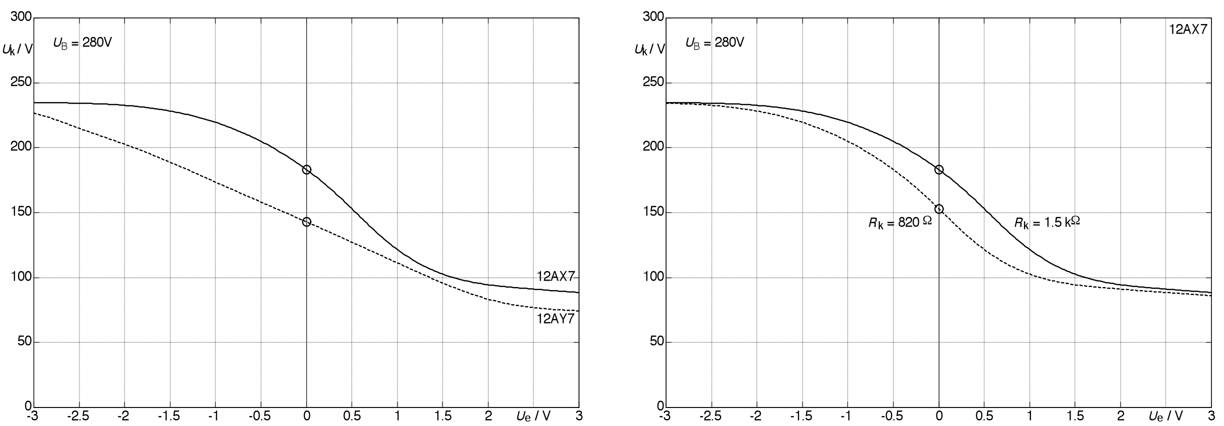


Fig. 10.2.6: Left: comparison 12AX7 vs. 12AY7 (1.5 kΩ // 25µF). Right: comparison 820Ω vs.1.5 kΩ (// 25µF). As in Fig. 10.2.4, the first tube was driven via a 100-kΩ-grid-resistor.

 Translated by Tilmann Zwicker

The first cathode-resistor (Fig. 10.2.7) determines the operating point of the first tube but the individual tube data also have significant influence. In **Fig. 10.2.7**, we see the results of measurements taken from several 12AX7 (Siemens, Valvo, Brimar, Mazda, Ultron, TAD). There are clear differences in the transmission characteristics as well as in the time-functions – this of course does dramatic effects on the level-dependencies of the harmonic distortion. Still, the attributes *good* or *bad* may be assigned with great caution only. Whether single-sided signal-limiting is preferred or objected to is a matter of taste, and the same holds for whether new or old tubes are utilized. A stringent correlation between tube data and tube age must not be expected – a clear correlation between tube price and **tube age** may be, though.

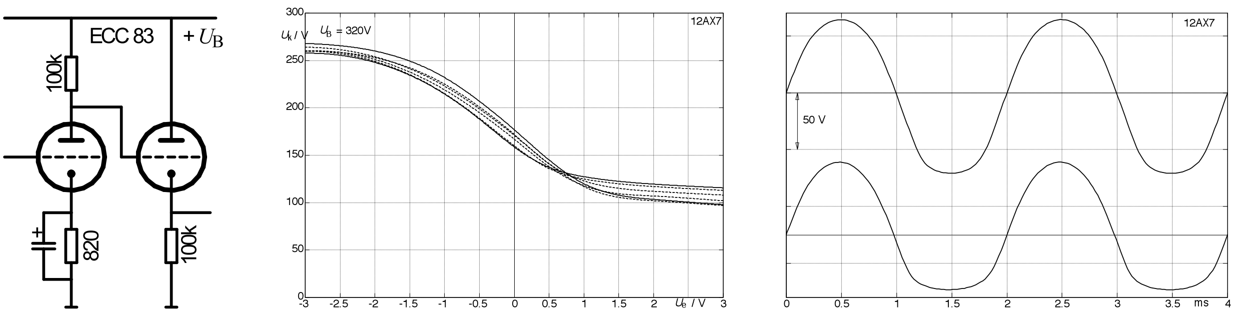


Fig. 10.2.7: Characteristics of 6 different 12AX7 tubes. Right: time functions (two different 12AX7).
First cathode-resistor = 820 Ω bridged with 25 µF; drive signal fed via a 100-kΩ-grid-resistor.

While we are on this subject: the opinion that **tubes produced back in the day** (NOS) are better, and consequently of course more expensive, holds only for the latter part. There might be something to the idea that the descendants of the old geniuses have plainly misplaced the recipes and do not know anymore how to build a high-quality tube. New tubes might have issues with microphonics, noise, a short lifespan, leaky seals, unsuitable getter♣, just to name a few criteria. But variations in the transconductance? The formula *higher transconductance = better* does certainly not work out, and a corresponding link to the price remains unclear, as well. The overdrive-behavior that is so important for guitar amplifiers is not specified in any data sheet for triodes, and, generally, neither is the grid current. A 12AX7 bought in 2008 may cost 6 € (advertised with tight bass, punchy mids and silky top end), or more than 13 € (tight bass, punchy mids and silky top end *with overall definition and brightness*). Or it could be priced at 25 € (great for warm clean tones and creamy overdrive). That is too expensive? Here is a 20-€-tube with "great warm clean tones and fat overdrive with smooth top end". Still not in your price-range? Hm … then maybe the 7-€-tube with "better gain and warm tone", or the 8-€-tube with "good gain, lots of treble and tight bass response"? Blimey – I've shelled out a 20-€-surcharge♥ for the tube-supplier scraping off the original labeling and replacing it by his company logo – shouldn't I be entitled a source to read up on the criteria that the tube (now knighted as "selected") will actually meet? Not a chance - "good gain", or "slightly better gain than Nr. 5" … that'll have to do. Or simply: "comes in the original RCA-boxing". That will set you back at least 30 €, though. But the real winner is: "*12AX7; enlarged grid giving a better articulation in the bass-range. The helix-shaped heating filament takes care of excellent noise-behavior and lowest microphonics*" – at no less that **42 € per piece**! Hopefully that extended bass-range is worth this kind of money-drain – given that the regular 12AX7 already extends down to 0 Hz. Of course, this sort of premium-stuff might be exactly what you were searching for forever. But then, the 5-€-no-name tube might have done the exact same trick. Faites vos jeux, ladies and gentlemen.

♣ materials that bind gas residues and improve the vacuum that way.

♥ Dear lawyers (including partners and colleagues in your firm scattered throughout the ROW): this is all just irreal satire. Ain't no spondu-licks coming through these tubes …

But now back to our actual topic: **Fig. 10.2.8** shows the harmonic distortion of the signals in Fig. 10.2.7. The differences between the left and right sections in Fig. 10.2.8 are due to just swapping tubes: take out the 12AX7 – plug in another 12AX7. Left, the 2nd-order distortion dominates up to -2.5 dBV; above that we see mainly 3rd-order distortion. Right, things are very different: 2nd-order distortion up to -11 dBV; from there on, about the same share for 2nd- and 3rd-order distortion. The closer the operating point gets to the end of the characteristic, the more dominant the 2nd-order distortion becomes for small drive levels. An ideal one-way rectifier (as an extreme example) would show only even-order distortion ($k_3 \equiv 0$).

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Fig. 10.2.8: Harmonic distortion of the signals of Fig. 10.2.7. 1st tube driven via a 100-kΩ -grid-resistor. Harmonic distortion attenuation $a_k = 20 \cdot \lg(1/k)$, k = harmonic distortion factor. Larger dB-values indicate smaller non-linear distortion. These figures are reserved for the printed version of this book.

Given such variances, wouldn't it be worth the while to use **selected tubes**, after all? That question is reason enough to check some offerings. A sample of 6 tubes sourced from a tube supplier was measured using the circuit seen in Fig. 10.2.7; the results are shown in **Fig. 10.2.9**. The small signal gain varies from v_U = 34.8 to 35.6 dB, and the operating points differ by as much as 20 V. The differences in the maximum and minimum achievable voltage are of similar magnitude, and thus in the symmetry of the curves, as well. "Asymmetry" would be the better term: in this circuit, this type of tube will be the source of pronounced single-sided distortion. Not that that's entirely undesirable in a Marshall amp … however, the precise reproduction of special distortion characteristics clearly is NOT warranted by the "selection" of tubes – as can easily be seen from **Fig. 10.2.12**. Except for the attribute "selected tube", no actual selection criteria are made public, and we can only speculate what the basis of the surcharge asked for these tubes could be. Maybe there is a selection for reduced microphonics – not entirely useless, but not a first priority in an intermediate amplifier stage, either.

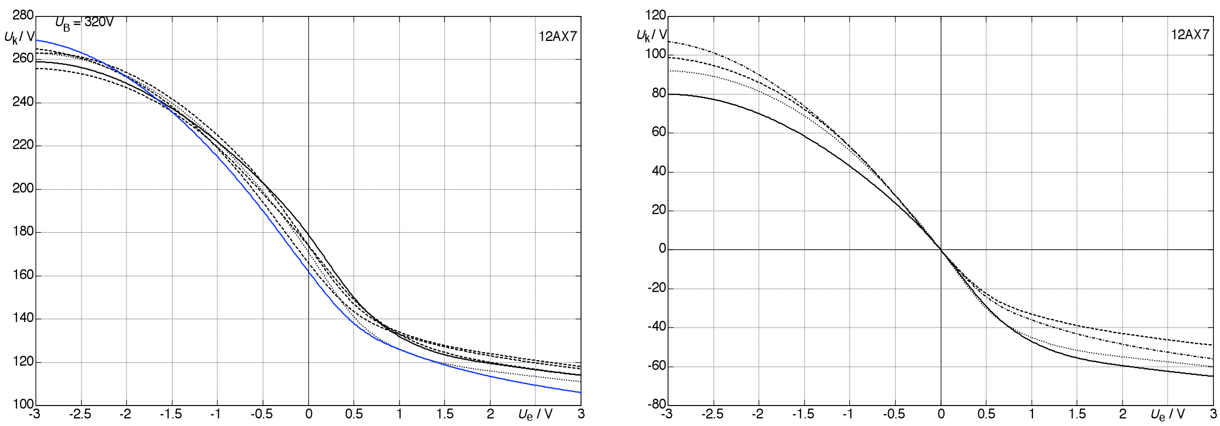

Fig. 10.2.9: Characteristic of 6 selected 12AX7 (supplier A); 4 of them in normalized presentation (right).

Fig. 10.2.10 shows measurements taken from 6 tubes provided by another supplier. The curves are indeed closer to each other although there are still differences in the details. Small signal gain is between 35.7 and 36 dB – a better match compared to our first example. The voltage limits, however, include a similar scatter so that we do not have a uniform distortion characteristic across several tubes, either (Fig. 10.2.12).

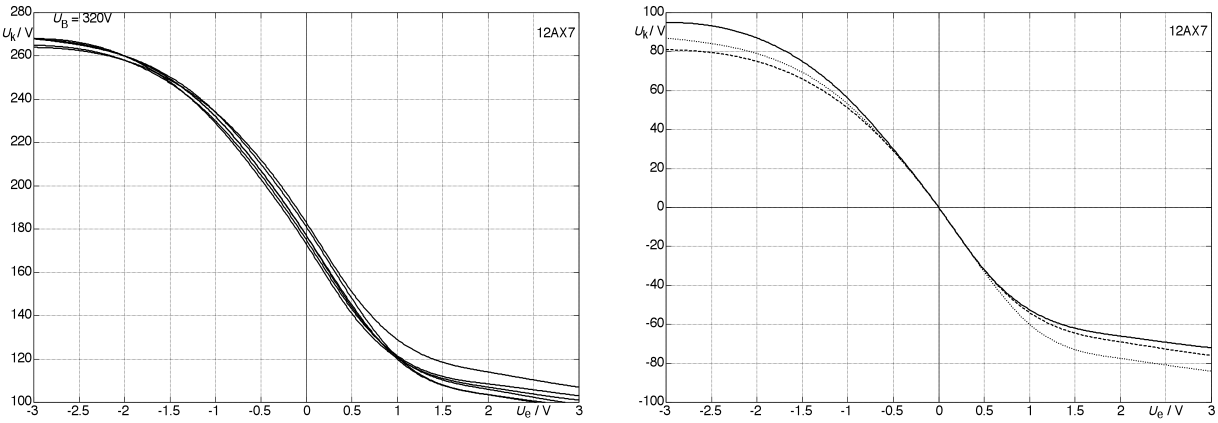

Fig. 10.2.10: Characteristic of 6 selected 12AX7 (supplier B); 4 of them in normalized presentation (right).

Last, let us take a look at 4 unselected tubes (all 4 from the same manufacturer), bought at a low price from a component discounter (**Fig. 10.2.11**). The small-signal gain v_U varies between 33.3 and 33.4 dB i.e. the gain factor this is 2 dB less than in the other samples. This can by no means be seen as a general deficit: whether the user prefers or dislikes the corresponding (small) reduction of distortion is a purely subjective rating.

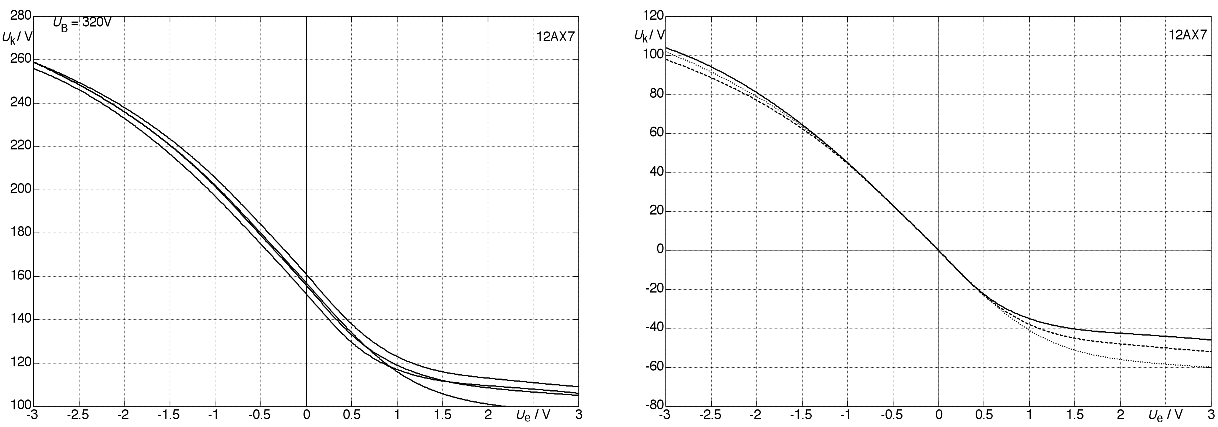

Fig. 10.2.11: Characteristic of 4 unselected 12AX7; 3 of them in normalized presentation (right).

In **Fig. 10.2.12**, again normalized transfer characteristics and harmonic distortion are brought face to face. The first sample of "selected" tubes shows measurable variance in the gain and – in particular – strong differences in the harmonic distortion; a common characteristic, however, cannot be established. The second and the third samples show a group-specific characteristic, but the variations within each group are still considerable – whether with or without "selection".

Of course, these measurements do not allow for the conclusion that *all* selected tubes offered on the market do not merit the term; the samples used here are too small for that. Still, inquiring about what the selection process in fact entails would appear to be highly advisable.

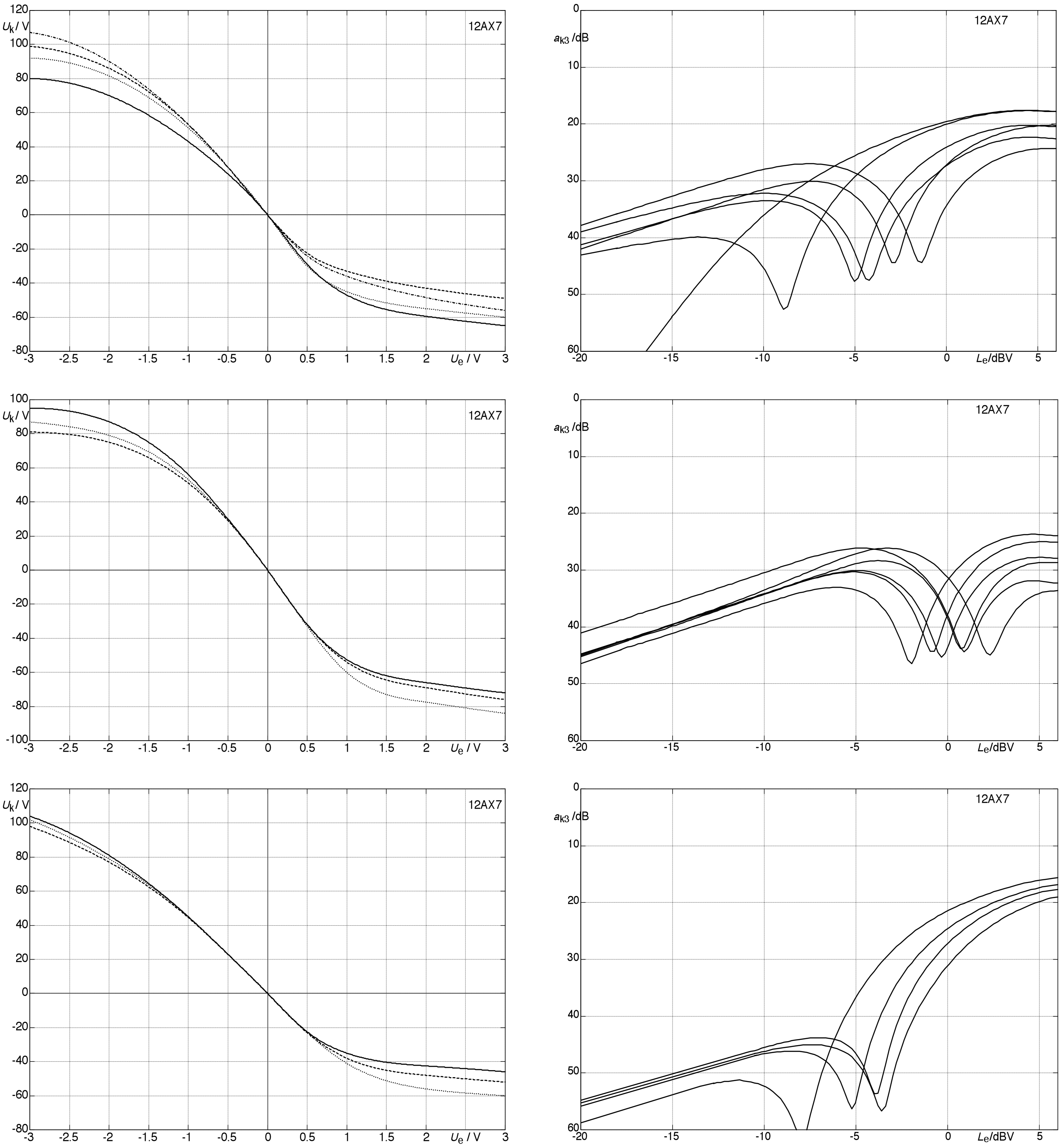

Fig. 10.2.12: Normalized transmission characteristics (left); harmonic distortion (right). Comp. Figs. 10.2.9-11.

Starting from the first Fender-circuits, the cathode-follower was subjected to two important changes until it arrived in Jim Marshalls JTM: 12AY7 → 12AX7, and 1500 Ω → 820 Ω. For the **VOX AC-30TB**, a third modification was added: the cathode-resistor at the cathode-follower tube was reduced from 100 kΩ to 56 kΩ, with the result that even without any drive signal, no less than 3 mA already flow through this tube. That is no laughing matter for a tube specified to carry 1,2 mA in its operating point. It won't be destroyed, but such a high current cannot be generated without the presence of a grid-current. This cathode-follower tube does not have a high-impedance input anymore but represents a non-linear load for the plate-circuit of the preceding tube. The latter is required to deliver a grid current of almost 1 mA which, considering that the plate resistor has a value of 100 kΩ, is no mean feat, and which will be the source of a special non-linearity. As is always the case with this special amplifier type, that might, however, not be generally undesirable.

Fig. 10.2.13 depicts the measurement results taken from the **VOX-circuit**: even without drive signal, the cathode-follower requires a grid-current of 185 µA. Measuring the differential input impedance (AC-resistance) of the cathode-follower resulted in the surprisingly small value of a mere 90 kΩ! This impedance-converter apparently does not feature the "extremely high" input-impedance typically found in such circuits, but is – due to its relatively high quiescent plate-current – even of quite low impedance. At high plate-voltages (U_{a1}), it loads down the preceding stage just like a 90-kΩ-resistor, and reduces the voltage gain of that stage by a quite sizeable 28%. With decreasing plate-voltage (U_{a1}), the input-impedance of the cathode-follower increases, after all; it therefore represents a non-linear load impedance. The transmission characteristic is strongly curved and the output-voltage swing is relatively small. This means that for large output voltages, the cathode-follower cannot provide enough current, and for small input voltages, the first tube is not sufficiently low-impedance. Not when using the 12AX7, anyway.

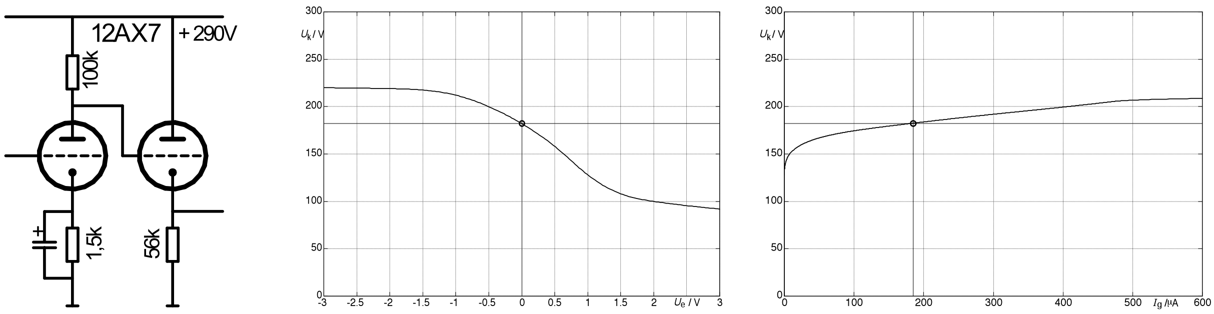

Fig. 10.2.13: Lefts: VOX AC-30TB. Middle: transmission characteristic of the overall circuit. For the measurement, the first tube is driven via R_{g1} = 100 kΩ. Right: grid-current of the cathode-follower tube.

Fig. 10.2.14 compares the summation- and the distortion-levels. The left-hand section shows the situation at the un-loaded 1st tube while the right-hand section describes the non-linear loading. The reduction of the summation level L_Σ by 2,8 dB and the growth of the distortion is clearly visible. Already at an input level of -15 dBV (178 mV), the 2nd harmonic (distortion) is a mere 30 dB below the level of the primary signal (i.e. k_2 = 3,2%). It will come as no surprise that the **internal impedance** (output impedance) of this cathode-follower is not at a by-the-book-value of 600 Ω but brings no less than 7 kΩ to the market: the operating point is not positioned by-the-book, either! Nevertheless: 7 kΩ are o.k. for the VOX-circuitry.

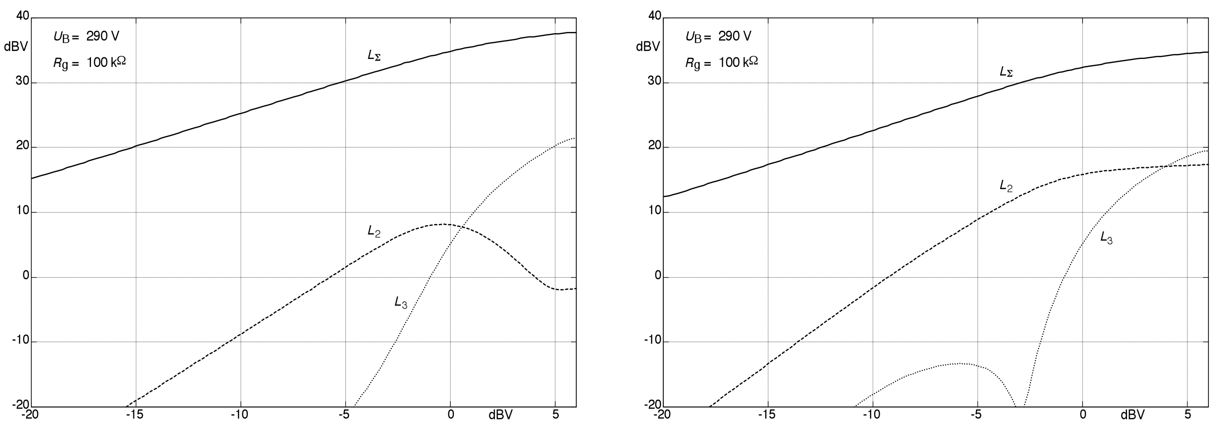

Fig. 10.2.14: Output-summation-level L_Σ, L_2 and L_3 of the VOX-circuit. Left: first half of the intermediate amplifier only (i.e. without cathode-follower). Right: complete circuit with cathode-follower.

The unusual selection of the operating point of the cathode-follower is the reason for strong 2nd-order distortion (k_2) showing up in the intermediate amplifier of the VOX. It is, however, difficult to surmise that there is any intentional design in this – the details too clearly fail to be reproducible. The non-linearity depends strongly on the supply-voltage, and on the individual tube in use, and it therefore appears – from one individual amp to the next - with varying distinction. (**Fig. 10.2.15**).

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Fig. 10.2.15: Level (left) and harmonic distortion (right) of the VOX intermediate amp; 8 different 12AX7. Grid-resistor in the first tube: R_{g1} = 100 kΩ. Supply voltage: U_B = 290V (compare to Fig. 10.2.13).
These figures are reserved for the printed version of this book.

All distortion measurements of the VOX intermediate amplifier were done with R_{k1} being bridged with a capacitor. During the history of the AC-30TB-circuit, there has, however, been a variant that fails to include this capacitor. With C_k = 25 µF, practically the whole relevant frequency-range receives an increase in gain of about 7.5 dB, while the 0.68-µF-capacitor found in some Marshall amps boost only the mids and highs (compare to Chapter 10.1). The treble-loss occurring upwards of 10 kHz happens in the first tube (R_{g1} plus Miller-effect). **Fig. 10.2.16** compares the frequency-responses measured with and without the cathode-capacitor.

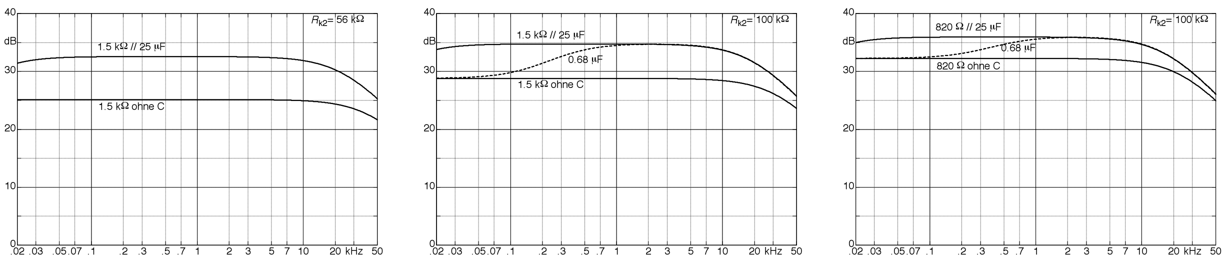


Fig. 10.2.16: Effect of the cathode-capacitor. In the VOX-circuit (left), the cathode-resistor is either bridged with 25 µF or left without a capacitor in parallel.

In the framework of discussing nonlinear distortion, the actual drive level is, obviously, of significance – there is no consistent benchmark for this, though. Guitar, playing style, setting of the tone- and volume-controls … all this determines the voltage arriving at the cathode-follower. Subtle playing may bring down the voltage level to below -20 dBV: in this case the non-linearity of the cathode-follower does not play any role. However, just turning up the volume control halfway generates – with a Stratocaster played in a normal way – easily voltage amplitudes of in excess of 1 V at the grid of the first tube in the two-tube-cathode-follower circuit. In particular the picking-attack will generate strong non-linear distortion in this scenario.

10.2.3 Mixing stage

Most guitar amplifiers feature more than a single "channel" i.e. there are several input jacks that are associated with different amplifier branches. These branches may vary in sound, in the distortion and/or in switchable effects. All branches are, however, fed to one and the same power amplifier, and this requires that the respective signals be added. Rather then the term "adding", the term "mixing" is often used – note that this does not refer to the process of the same name used in RF-engineering and designating circuits for frequency conversion. For the present context, we mean: **mixing = adding**.

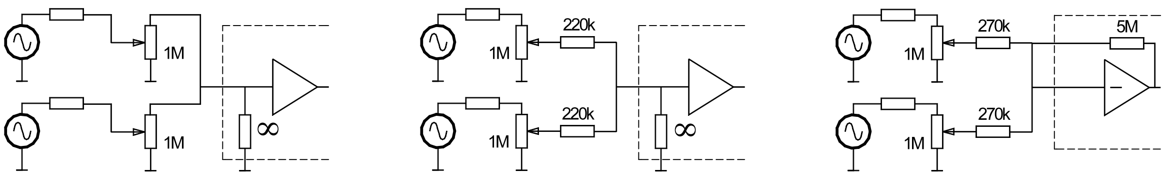


Fig. 10.2.17: Circuit concepts for signal addition: reverse-mode, standard-mode, active-mode (left to right).

Three often-implemented circuit concepts are shown in **Fig. 10.2.17**. The so-called reverse-mode was often found in early amplifiers; it was soon replaced by the standard-mode. Passive circuitry has the general disadvantage that the potentiometers influence each other: if the volume control in one channel is fully up (α = 100%), and if the second volume control is now also turned up (β = 100%), the gain factor of the first channel can be reduced by up to 6 dB because of the mutual loading between the two channels. **Fig. 10.2.18** shows this influence dependent on the center-tap position of the respective other potentiometer (β).

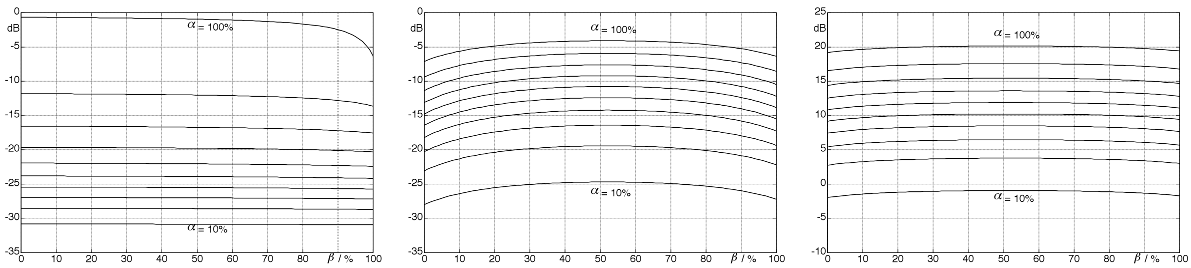


Fig. 10.2.18: Mutual influence of the two potentiometers; α = CH1, β = CH2. Figures assigned as in Fig. 10.2.17. Potentiometer = 1 MΩ, mixing resistors = 220 kΩ and 270 kΩ, respectively. Passive modes: gain up to the summation point. Active mode: gain incl. tube stage (v = -50).

The internal impedance of the sources (amounting to about 40 kΩ for triode-amplifier stages in common-cathode configuration: tube // plate-resistor) has an effect on the "counter-side" as the potentiometers are turned up, and attenuates the "other" signal. Additional summation-resistors (in series with the potentiometer center-tap) reduce this effect for the **standard-mode**. In the Fender Deluxe 6G3, for example, we see 220-kΩ-resistors at this point in the circuit, but there are also amps that use 470 kΩ (e.g. the Bassman 6G6). Larger summation-resistors give a higher-degree independence of the controls but do have the disadvantage that noise is likely to increase, and that the treble-response will probably get worse. In the third variant, the **active-mode**, a negative-feedback-resistor reduces the gain as well as the input impedance (current-voltage-feedback). Given high open-loop gain and strong feedback, the contra-lateral influence can be practically eliminated. A small dependency remains in the typical tube amp with v = – (30 ... 50) but this is practice is of no bother. As another effect of the negative feedback, maximum gain and harmonic distortion decrease.

Active mixing-stages are not often seen in guitar amplifiers: they surfaced in Fender amps in the mid-1950's (5E4, 5E5-A, 5D6-A) but disappeared again shortly afterwards. The standard-mode is by far the most often used, with mixing resistors of 220 – 470 kΩ. Moderately reducing the mixing resistors does not bring much advantage regarding the gain but increases the upper cut-off frequency (while deteriorating the mutual interaction). With the potentiometer center-tap positioned mid-way, the source-impedance that the following tube-grid "sees" is about (P/4 + R)/2, with P = potentiometer-resistance and R = mixing resistor. Typical values of this source impedance are found to be in the region of 250 kΩ. In conjunction with the tube-input-capacitance (up to 150 pF due to the Miller-effect), a 1st-order **low-pass** with a cutoff-frequency of 4 – 8 kHz results. In some amplifiers, the corresponding slight treble-loss is counteracted via a **bridging-capacitor** that bridges potentiometer and/or mixing resistor. This may be implemented only in one of the two channels because otherwise the effect would suffer. Manufacturers like to designate the channel modified that way with terms such as "Bright" or "Treble or "Instrument", while the other channel is dubbed "Standard" or "Normal".

In **Marshall's JTM-45**, a guitar amplifier from the early 1960's, the signal addition is done via two 270-kΩ-resistors in the beginning – just like in the Fender the JTM was modeled after. Soon, however, there is a change to 470-kΩ-resistors; these remain for several model generations. To compensate the associated treble-loss, bridging capacitors with model-specific value-variations are installed. The early Marshall amps were available in versions for guitar (lead), for organ, for bass and for use as PA, with the technical distinction between them mainly being the differing values of the bridging capacitors and the mixing resistors.

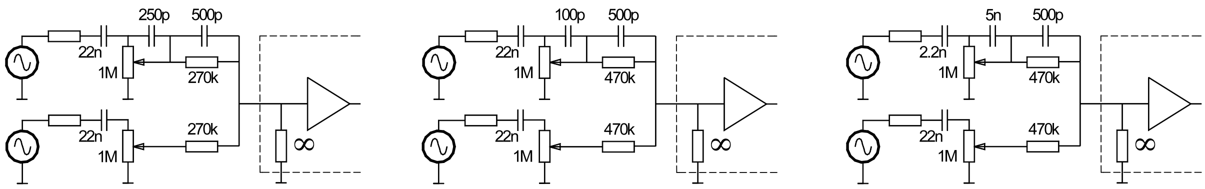


Fig. 10.2.19: Marshall-amplifier, adding stages with different-value components.

Fig. 10.2.19 shows three versions of the mixing stage; for the first (on the left), **Fig. 10.2.20** indicates the frequency-responses for different positions of the respective volume-control. The grey areas depict the ranges of mutual influence of the two controls. Depending on one's position in the hierarchy of Marshall-ites, these results may be interpreted as testimony to genius manifoldness, or as ghastly circuitry-botch-up.

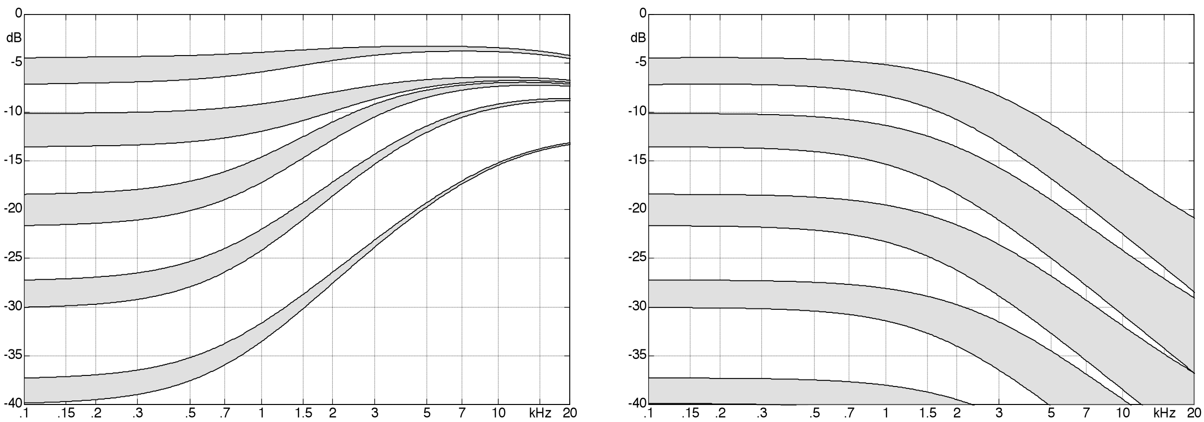


Fig. 10.2.20: Marshall JTM-45, mixing stage. Left: frequency-response of the "High Treble" channel, right: "Normal"-channel. The grey areas show the mutual influence between the two volume-pots.

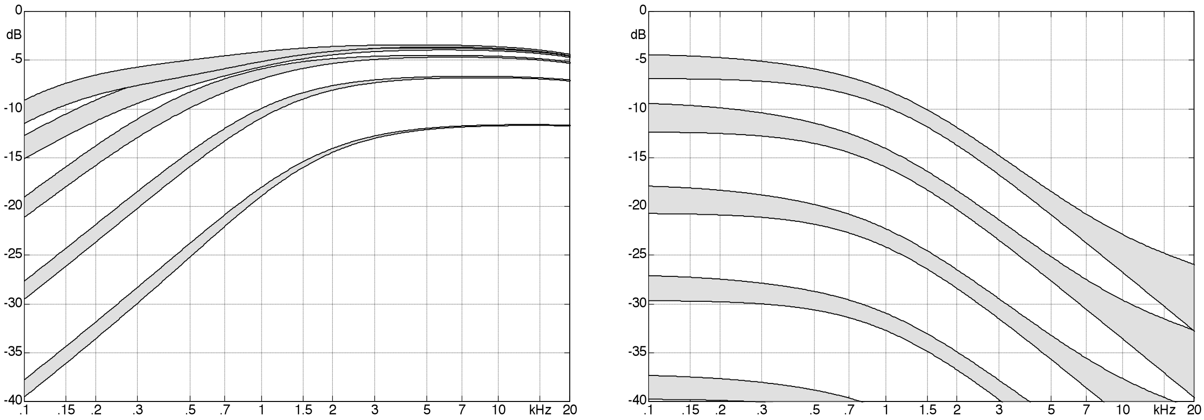
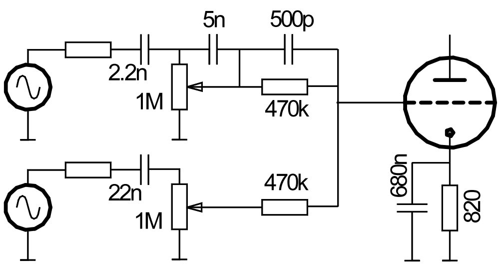


Fig. 10.2.21: Marshall Type 1987, mixing stage. Left: frequency-response of the "High Treble" channel, right: "Normal"-channel. The grey areas show the mutual influence between the two volume-pots

In **Abb. 10.2.21** we see the frequency-responses of the circuit shown on the right in Fig. 10.2.19. The change to the unusually large 5-nF-capacitor results in a special low-cut. Also, in the upper range of the volume control (i.e. where the user usually "lives"), it operates almost solely as an adjustable bass-cut. That is quite successful, as one can hear. The reduction of the coupling capacitor to 2.2 nF makes for an additional low-cut. Since apparently the sound was *still* not aggressive enough, the cathode-resistor was bridged not (as Fender would have it) with a large electrolytic capacitor, but with a 680-nF-capacitor (**Fig. 10.2.22**) that makes this stage run at maximum gain only for higher frequencies. At low frequencies, there is a slight negative feedback. Some Marshall amps had a further capacitor to bridge the cathode-resistor in the pre-amplifier, other completely dispensed with these caps. There is, after all, neither "the" Marshall-circuit nor "the" Marshall-sound.



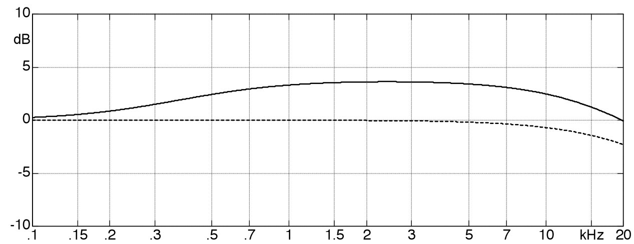


Fig. 10.2.22: Left: cathode-resistor bridged by a capacitor in the Marshall amp types 1987 and 1959. The right-hand picture shows the treble boost resulting from the cathode-capacitor.

10.3 Tone-Controls

Just to state it right upfront: the secret of a great-sounding guitar amp does not lie in its tone controls (tone-filter). Of course, these modules are necessary to adjust bass, middle and treble to the subjective desires, but modifications to the tone controls normally will not convert a bad amp into a great one.

The first guitar amps often had merely a simple treble-control. Fender's Champ even had only one solitary knob: *Volume*. If sound variations were indeed indispensable, you had to do them on the guitar. The Deluxe at least already had a treble-control, and over the years, further controls were added. In the 1950's, your standard helping of tone-control included a Bass and a Treble-knob, and later some chosen few received a middle-control in addition. Marshall copies Fender's tone-control circuit (with minor modifications), and in Jennings' VOX-amps, a comparable filter-stage is found. And there you have it: the glorious Big Three – most subjectively chosen, of course. Trying to put together even only an approximately representative selection of all tone-controls developed over the years would go WAY beyond the scope planned here, and so we will limit ourselves to a only few circuits.

Set to their middle ("neutral") position, the tone controls in a HiFi-amplifier need to give a frequency-independent reproduction. The tone controls in a guitar amplifier do not have to perform that way, because the amp is – together with the loudspeaker – still a part of the sound generator and contributes to the sound. Although the tone controls may include frequency-selective filtering of more than 20 dB, it is not the only filter-stage in a guitar amp. The input capacitances of the tubes have (in conjunction with the usually high-impedance circuitry) the effect of a treble-cut. Bridging capacitors (over-) compensate this via a treble-boost. Intentionally small coupling capacitors attenuate the lows, as do small cathode-capacitors. Frequency-selective negative feedback in the power stages yields brilliance, output transformers may contribute resonance-accentuations and/or bass-cuts, and at the end of the transmission chain we have the loudspeaker with its only weakly dampened resonances. No, this transmission is everything but frequency independent – and that is what makes it so desirable.

10.3.1 Bass-Middle-Treble

As an example for a **passive tone control** we chose a circuit that is included in many Fender-amps, but (in more or less modified versions) also has found its way into amps by Ampeg, Kitty Hawk, Marshall, Mesa Boogie, Music Man, Randall, Rickenbacker, Roland, Selmer, Solton, VOX, and many more. The term "passive tone control" indicates that the frequency-dependent filtering is done exclusively via passive components, i.e. by resistors and capacitors. The tube-stages grouped around the tone control contribute frequency-independent gain. As an approximation, we may ignore for now that this is not fully correct. In an *active* tone control, the RC-network is integrated into the feedback loop of a tube, and corresponding circuits have a significantly different structure. Fundamentally, inductances also count as passive components – but they are not liked, due to their relatively large build. At the most they are included as exotic birds, if at all.

 Translated by Tilmann Zwicker

In **Fig. 10.3.1**, we see a good example for a simple passive tone-filter. This circuit was deployed in early Fender-amplifiers (e.g. the 5E4) but may be found in variations also in radios and similar devices. Turning down the bass control (i.e. moving the tap in the figure fully to the right) results in a readily comprehensible situation. What remains now is merely a complex-valued voltage divider that can be further simplified if we take the load resistance as infinite. The current becomes independent of the position of the center-tap, and depends on the frequency only as a 1st-order function (cutoff frequency = 653 Hz), despite the presence of *two* storage-elements. The output voltage, as multiplication of this current with the transverse impedance, also merely has a 1st-order dependency on $p = j\omega$, and an appropriate adjustment of the treble control even results in a 0-order-system with frequency independent transmission (32.2 dB attenuation). The right-hand diagram in Fig. 10.3.1 shows the transmission functions of the divider without load; the position of the center-tap is the parameter.

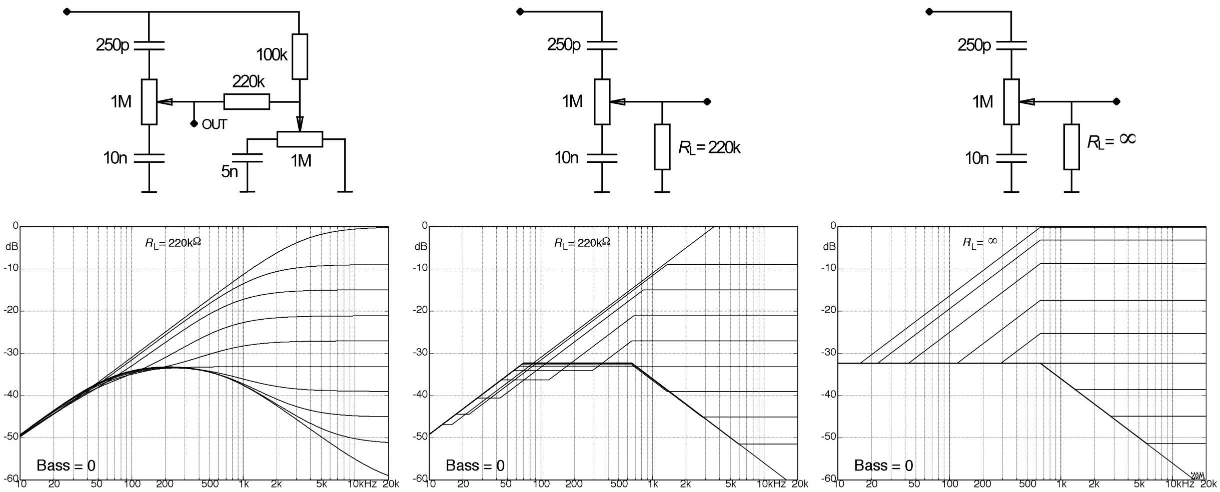


Fig. 10.3.1: Simple treble-filter. The circuit on the left was used in the 5E4 Super-Amp; the circuits to the right are simplifications for the bass-control turned down. See also Fig. 10.3.3.

Introducing a load-impedance yields a 2nd-order transmission-function that, as an approximation, can be seen as load-less divider with an additional high-pass (f_g = 70 Hz). The middle picture shows this scenario as a Bode-diagram with approximation lines. In the left-hand picture, we see the complete magnitude-frequency-response. In a real circuit, it will be necessary to consider the input capacitance of the subsequent tube; this capacitance can easily amount up to 100 pF due to the Miller-effect. The resulting minor treble-attenuation is only felt above about 10 kHz. **Fig. 10.3.2** shows a peculiarity of the Fender-circuit that sets it apart from the tone-filters usually found in audio-engineering: while the latter keep the cutoff frequency constant and fan out the curves symmetrically, the cutoff-frequency for the Fender-filter changes as the treble control is adjusted.

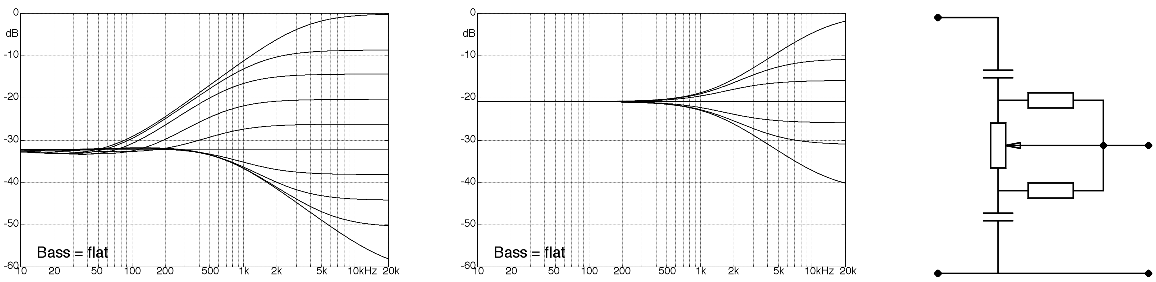


Fig. 10.3.2: Comparison of magnitude-frequency-responses: guitar amp (left), audio amplifier (middle and right).

Fig. 10.3.3 depicts the effect of the Fender tone-filter in 6 diagrams. Again, pronounced differences compared to classic audio-filters are apparent: treble- and bass-attenuation influence each other, and for bass and treble fully turned up, a rather selective mid-cut results. The latter is a specialty that will remain in almost all later Fender amplifiers.

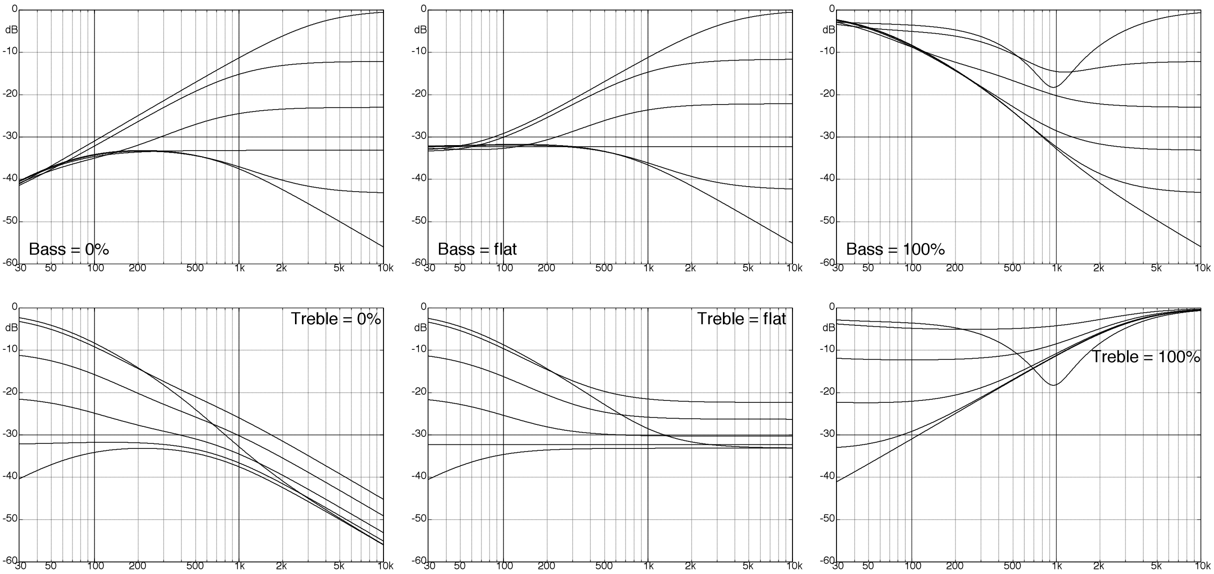


Fig. 10.3.3: Frequency-responses of the Filter circuit acc. to Fig. 10.3.1 (Fender Super-Amp 5E4, ca. 1955)

The structure of this tone-filter has some similarities to the mixing-stage discussed in Chapter 10.2.3: treble and bass are divided up into two parallel channels, then high- and low-pass filtered, respectively, and finally added up again at the output. The 5-nF-capacitor shorts high frequencies to ground; as such it has a function similar to that of the 10-nF-capacitor. Combined with a desire to cut cost, it was presumably this similarity that led to a merging of the two capacitor-branches. To keep the effect of the Treble filter when the Bass-control was turned down, a resistor was required between the 10-nF-capacitor and ground … and you got a tone-filter that makes do with **only two capacitors** (**Fig. 10.3.4**).

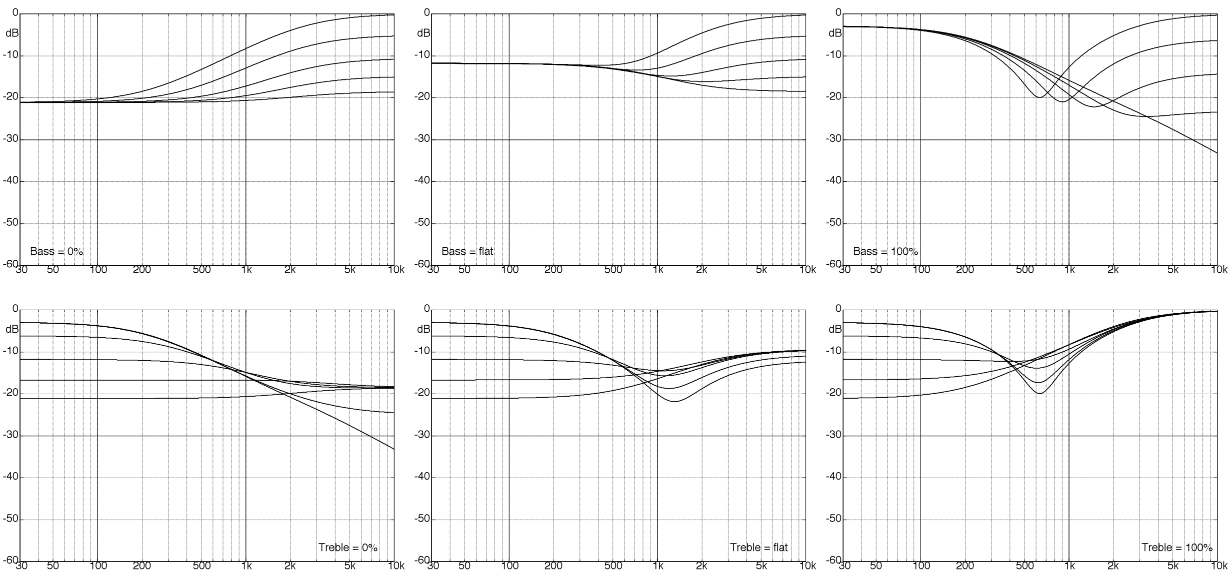


Fig. 10.3.4: Frequency-responses: tone-filter of the Super-Amp 6G4. Circuit given in Fig. 10.3.8.

Supposedly, the control options of this simple filter were seen as too limited, after all, because very soon there was the revision 6G4-A (**Fig. 10.3.5**): an updated filter-circuit with no less than 4 capacitors, and with a special treble-potentiometer sporting an additional **tap**. Apparently, this development was worth the effort since the Tremolux (6G9) received it as well, and since it was also used in the Bandmaster (6G7-A) and the Vibrolux (6G11), albeit with small component modifications in the latter two.

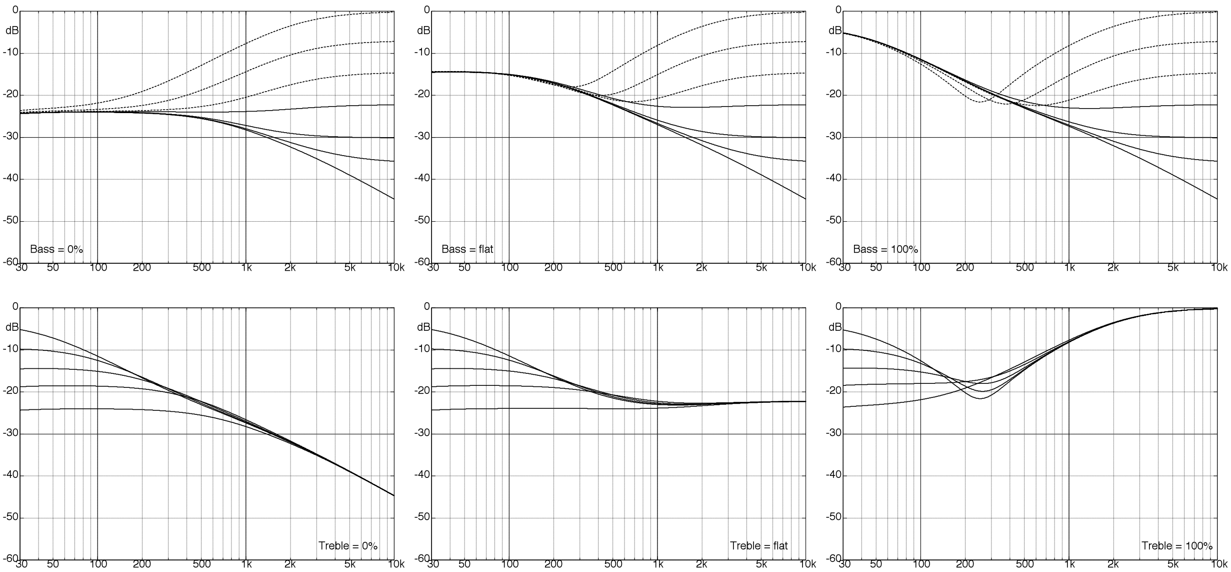


Fig. 10.3.5: Frequency-responses: tone-filter of the 6G4-A. Circuit as in Fig. 10.3.8.

Nevertheless, the pot with the special tap disappeared again already in the following amplifier generation, and around 1963 a circuit was developed that would go down in history as the mother of all tone-filters – to be found in this or very similar configurations in VOX, Marshall and many other guitar amplifiers (**Fig. 10.3.6**). In fact, the range of settings is not that big, but it apparently fits the combination Fender-guitar + Fender-amplifier perfectly. The individual component values are subject to variations at Fender as well as for the many copycats (in particular the "mid-scoop" is shifted back and forth in its frequency position), but the basic topology is now set.

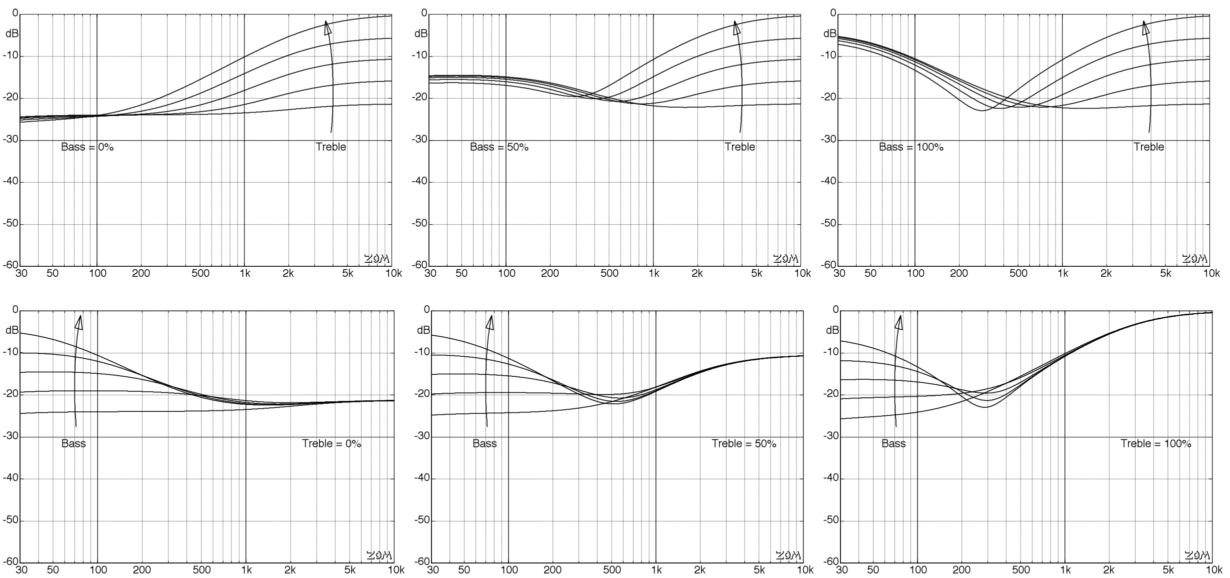


Fig. 10.3.6: Frequency-responses: tone-filter at he beginning of the 1960's. Circuit as in Fig. 10.3.8.

Also, the new filter circuit (AA763, Fig. 10.3.8) allows for the addition of a **Middle-Control** in addition to Bass- and Treble-Controls. The required effort is rather small: the fixed 6.8-kΩ-resistor of the first version is simply replaced by a 10-kΩ-potentiometer.

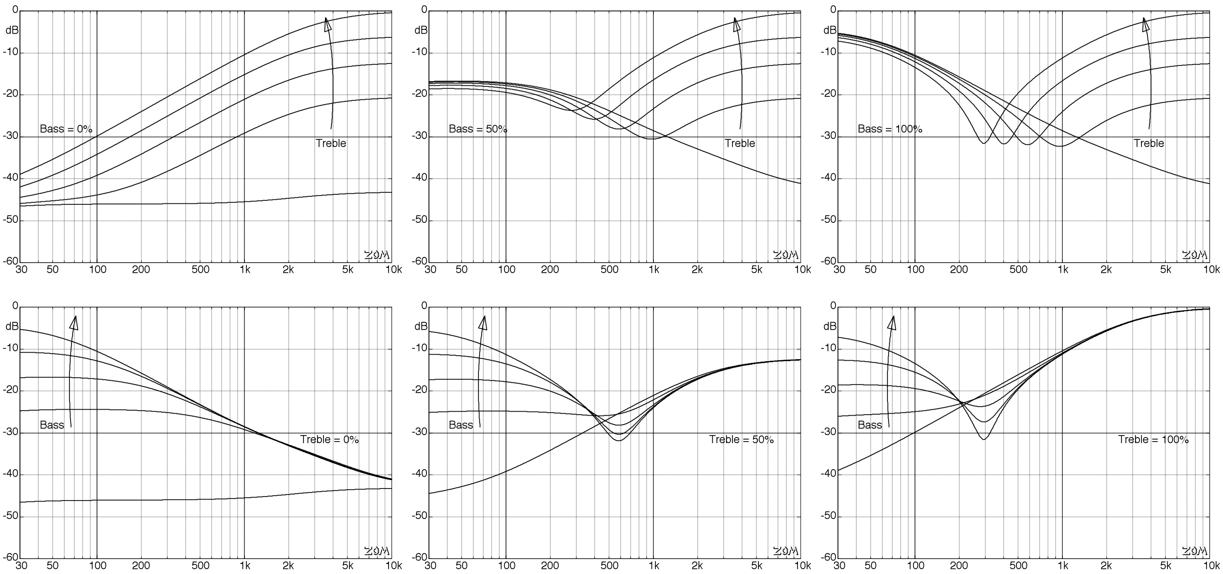


Fig. 10.3.7: Frequency-responses of the tone-filter with middle-control (R_M = 500Ω). Compare to Fig. 10.3.6.

Fig. 10.3.8 documents the development of the Fender tone-circuit. The number of capacitors changes from two to four until a simple 3-capacitor-circuit is found the topology of which to this day is seen as a standard.

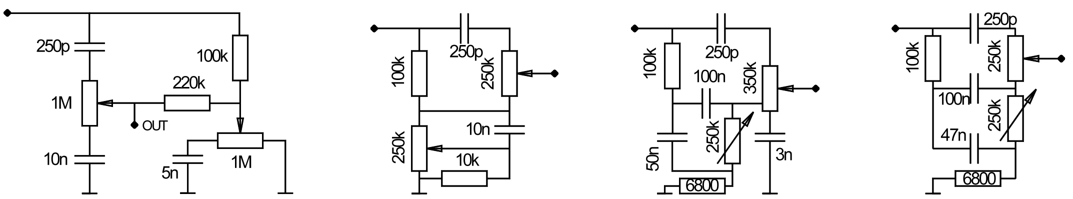


Fig. 10.3.8: Fender tone-filter circuits: 5E4, 6G4, 6G4-A, AA763 (left to right).

As mentioned, there were changes now and again in the values of the components of the AA763-tone-filter: apart from the variations on the 6800-Ω-resistor (middle-pot), the 47-nF-capcitor was subject to several modifications and varied from 22 nF to 33 nF and on to 47 nF. The effect of this change in capacitance is shown in **Fig. 10.3.9**: if the Bass control is not entirely turned down, the spectral components below 500 Hz are boosted by the reduction of the capacitance. With the bass-pot at "0" nothing changes because in the relevant frequency-range the parallel connection with the 100-nF-capacitor acts approximately as a short compared to the 100-kΩ-resistor. This holds for 22 nF as well as for 47 nF. It is difficult to find a clear criterion for the choice of this capacitor-value in Fender amps. Some amplifiers such as the **Showman** or the **Twin** start with 47 nF in 1963 and keep that value. The **Bandmaster** receives the 47-nF-capacitor in 1963 but 5 years later this is changed to 22 nF. The **Pro-Amp** sports a 33-nF-capacitor to begin with (AA763), but that is changed to 47 nF in the same year (AB763) – and 6 years later we find a 22-nF-capacitor. In the **Super-Amp**, the capacitor-history is different: it starts out with 33 nF (AA763), then in the same year sees the change to 22 nF. Yet another approach in the **Deluxe**: 33 nF in the AA763 and the change to 47 nF in the same year. Must be magic …

Fig. 10.3.9: Differences between 22 nF (fine line) and 47 nF (heavy line) in Fender tone-filters; R_M = 6800Ω.
These figures are reserved for the printed version of the book.

The frequency-responses shown so far do not consider the peripheral circuitry. The source impedance of the preceding stage and the input impedance of the subsequent stage change the curves – but not fundamentally; in fact the differences are rather marginal. In Fender amplifiers, the source impedance typically amounts to 30 – 40 kΩ, which is low enough that we approximately have a stiff voltage source. The load of the tone-circuit is either a high-impedance tube-input or the volume-potentiometer. The latter is at 1 MΩ (rarely also 500 kΩ) of sufficiently high impedance; the output of the tone circuit therefore can be seen as operating without load. For the uppermost frequency-range, however, we do need to consider the **input capacitance** of the subsequent tube. Due to the Miller-effect this has to be assumed to be 100 – 150 pF, leading – in conjunction with 250 kΩ (see below) – to a cutoff frequency of 6.4 or 4.2 kHz, respectively. The corresponding loss in brilliance makes itself felt most when the center-tap of the volume potentiometer is set mid-way, because here the internal impedance of the pot is largest (R/4 = e.g. 250 kΩ). In order to counteract the treble-loss, already the first guitar amps had a **bright-capacitor** installed that bridged the upper part of the volume pot. In its left-hand section, **Fig. 10.3.10** shows the treble-loss due to the capacitance, and in the right-hand section the effect of the bright-capacitor. The figure focuses on the transmission characteristic given by source impedance (38 kΩ), volume pot (1 MΩ), bright-capacitor (120 pF), and input capacitance (150 pF); the additional attenuation of the tone-filter is not shown to maintain a straightforward display.

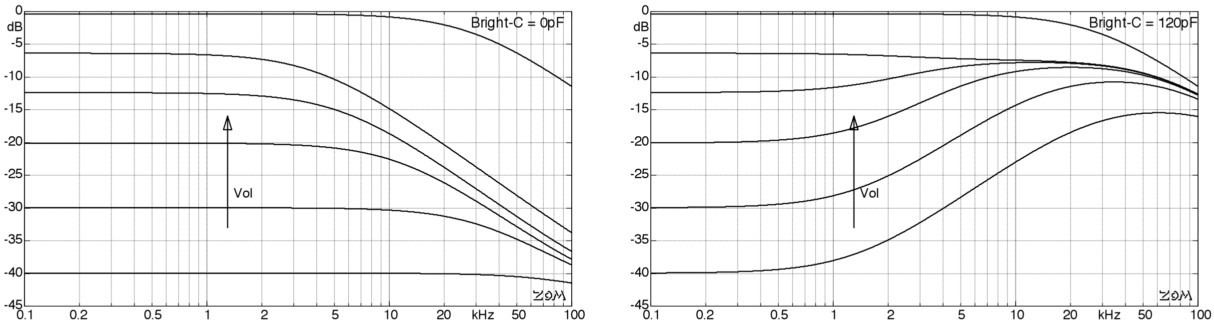


Fig. 10.3.10: Treble-loss due to capacitive loading of the volume pot (left), treble boost via bright-C (right)
Source impedance R_Q = 38kΩ, 1-MΩ-potentiometer, input capacitance of the subsequent stage: 150 pF.

The Fender tone-filter designated AA763 is again shown in **Fig. 10.3.11**, this time in comparison to two competitors that originated at approximately the same time: the VOX AC-30TB and the Marshall JTM-45. The basic structure is identical but there are characteristic variations in the details. For example, the Fender filter shuts the signal off completely with all controls turned down fully – the other filters avoid this awkward property. The individual component values differ substantially so that three distinct circuits emerged, after all – despite all similarities.

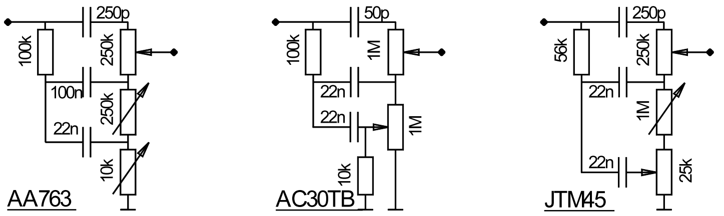


Fig. 10.3.11: Comparison of tone-filter circuits: Fender, VOX, Marshall. The filters are loaded differently: high impedance for Fender and MARSHALL, 360 kΩ for VOX (Miller capacitance to be added to each).

In **Fig. 10.3.12** we see the transmission characteristics of the **VOX**-Filter (AC-30TB). The low-cut is particularly conspicuous; it is due to an RC high-pass not shown in the figure. The Marshall-filter (**Fig. 10.3.13**) is different, again: the aim here apparently was a small attenuation of the filter stage. *(Translator's note: incidentally, this Marshall-tone-circuit is a direct copy of the circuit found in the last tweed Fender Bassman 5F6-A that had – in the tone-control-department – similar advantages and disadvantages.)* This attenuation is further reduced in the subsequent versions of the amplifier (JTM-50, **Fig. 10.3.14**) by replacing the 56-kΩ-resistor by 33 kΩ and the 250-pF-capacitor by 500 pF.

Fig. 10.3.12: Frequency-responses: the tone-filter of the VOX AC-30TB (incl. 580-Hz-high-pass).
These figures are reserved for the printed version of this book.

The **differences in tone** of the three amplifiers under scrutiny here are, however, not principally based on the different filter circuits. Only several stages cooperating make for the individual sound. For example, the high-impedance power-amp output of the AC-30TB results in a strong bass-boost (Chapter 10.5.7) that is found in Fenders only to a much smaller degree. Marshall amps, on the other hand, offer the presence filter integrated into the power amplifier stage; it brings a special treble-boost that the VOX lacks. We find further differences in the overdrive-behavior and in the loudspeakers used: the latter typically work in an open combo-cabinet in the Fender and VOX amps – for the Marshall, however, the bass-heavy 4x12-enclosure is employed. While the tone-filter is a substantial part of the overall system, its respective special realization should not be credited with any exaggerated importance.

Fig. 10.3.13: Marshall JTM-45. The Treble-boost from preceding stages is not considered.
These figures are reserved for the printed version of this book.

Fig. 10.3.14: Marshall JTM-50. The Treble-boost from preceding stages is not considered.
These figures are reserved for the printed version of this book.

The following examples show that, in tone-filters, “more” is not necessarily “better”: in the Fender circuit we find two to four capacitors but the Sound-City-filter has six of them! Or even 10, as shown in **Fig. 10.3.15**. Not bad, but short lived. If this filter structure were superior, it would have asserted itself in products by the competition, as well – but that didn’t happen, and the circuits disappeared again from the market.

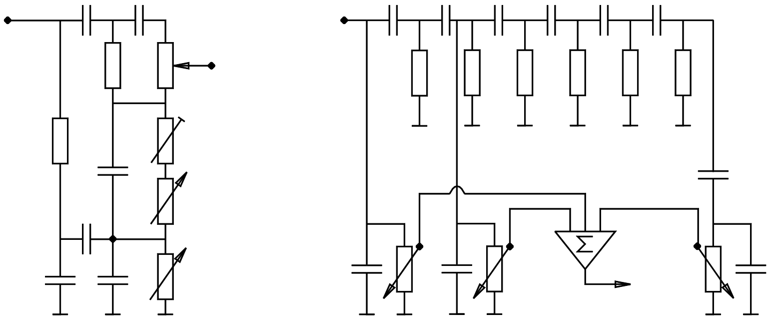


Fig. 10.3.15: tone-filters in Sound-City amps:
Left: CS100B.
Right: L/B 120 Mark IV.

Simple tone-filters do not stand in the way of creating a convincing amplifier, as the **Marshall 18-Watt-amp** (examined in the following) proves. This amplifier was produced from 1965 – 1967 and has a lot of fans despite its rather spartan filter-network. In the “Normal”-channel we find a single tone control: cut either treble, or bass – that’s it. Similarly, there is only one simple Tone-knob in the “Tremolo”-channel: more treble or less treble, interactively coupled to the volume-pot.

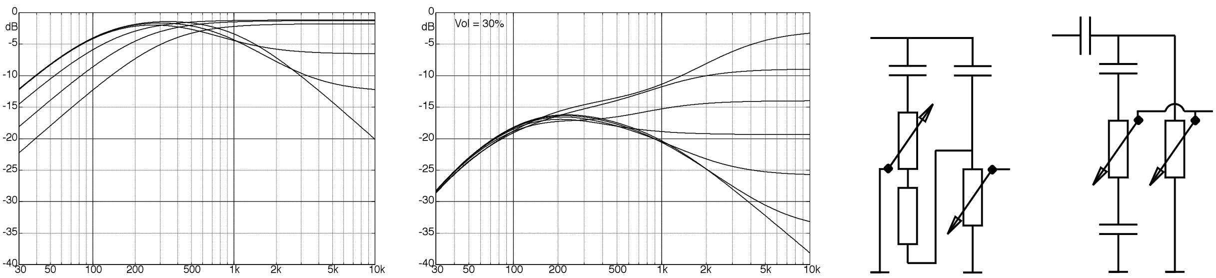


Fig.10.3.16: Frequency-response of the Marshall 18-W-amp; left: “Normal” channel; right: “Tremolo” channel.

Very similar circuit concepts are found already 10 years earlier in the Fender “Deluxe-Amp” amplifier; the volume-pot is merely connected “in reverse” to facilitate the connection of a second channel. Even today, these very simple old amps are not at all “out” but enjoy cult-status in the use for club-gigs or in the studio. Very obviously, a complicated tone-filter is not necessary to amplify an electric guitar. Question to Lenny Kravitz♣: *"How do you get this tone?“* Answer: *"Well, you just plug an Epiphone into a Tweed Deluxe, crank it to 10 ... and that’s it.”*

On the other end of the spectrum of complexity we find amplifiers that offer almost infinite variability using multi-band graphical and/or parametric equalizers (Chapter 10.3.2). They are predestinated for the creation of very “different” sounds, but the majority of guitar players seem to be able to do without them.

♣ Gitarre&Bass 06/04

10.3.2 Equalizer (EQ)

A filter that allows for narrow-band changes in the spectrum (or in the transmission function) is called an equalizer. Besides a basic gain that we assume to be 1 ($\hat{=}\,0\text{dB}$) in the following, there are 3 parameters that define the transmission behavior of an equalizer: center-frequency, boost and Q-factor (**Fig. 10.3.17**) The center-frequency f_x is the **frequency** at which the gain assumes it maximum (or minimum) value, the **boost** β specifies the gain at f_x, and **Q-factor** Q determines the bandwidth. For a so-called parametric equalizer (EQ), all three parameters are adjustable while for a so-called graphic EQ, only β is variable, with f_x and Q fixed at predetermined values.

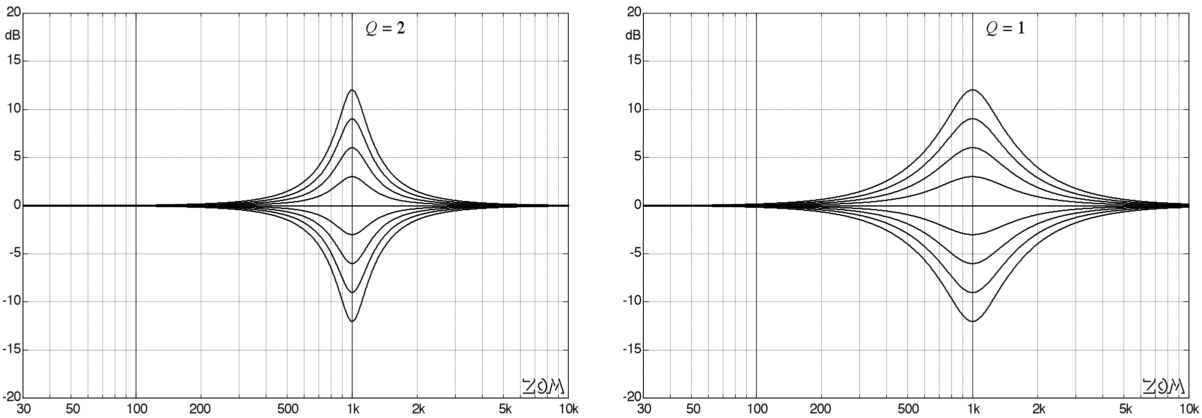


Fig. 10.3.17: Equalizer characteristic. $B = 20\cdot\lg(\beta)$ = [-12 -9 -6 -3 0 3 6 9 12]dB, f_x = 1 kHz.

In Fig. 10.3.17 we see two different groups of curves. f_x and B are self-explanatory, but the Q-factor requires some supplementary comments. Often, the Q-factor is determined from the relative bandwidth measured as the distance of the -3-dB-points on the graph. This definition is, however, useless for an EQ e.g. because for a 2 dB-boost no -3-dB-points can be defined at all. The correct definition results from the transmission function $\underline{H}$:

$$\underline{H} = \frac{\omega_x^2 + p\cdot\omega_x / Q_Z + p^2}{\omega_x^2 + p\cdot\omega_x / Q_N + p^2} \qquad p = j\cdot 2\pi f \qquad \omega_x = 2\pi f_x$$

As can be seen, this filter has a pole-Q-factor Q_N and a zero-Q-factor Q_Z. For $f = f_x$, we get $b = Q_N\ /\ Q_Z$. In order to define *one single* Q-factor for an equalizer, an infinite number of possibilities present themselves; customary are two (different!) definitions. Either we keep the denominator-Q-factor constant and vary the boost-factor via the numerator-Q-factor; this filter-type is called **constant-*Q*-equalizer**, and the denominator-Q-factor is specified as the Q-factor of the equalizer. Or we link numerator- and denominator-Q-factors via $Q_Z = Q/\sqrt{\beta}$ and $Q_N = Q\cdot\sqrt{\beta}$; in this case we specify as Q-factor of the equalizer: $Q = \sqrt{Q_N \cdot Q_Z}$. Connecting two equalizer of the second variety in series with f_x and Q correspondingly identical in both EQs, and the boost-factors set reciprocally ($\beta_1 = 1/\beta_2$), the effects of these two equalizers compensate each other completely. They are inverse to each other, and therefore this EQ-type is also called **inverse EQ** (the filter shown in Fig. 10.3.17 is of this type). For the constant-*Q*-equalizer, however, a corresponding series-connection does not lead to a complete compensation: the attenuation is of a smaller bandwidth than the amplification (**Fig.10.3.18**). These differences (if they are of any importance at all) play a role only for graphic EQs, because all parameters can be freely adjusted in the parametric EQ, anyway.

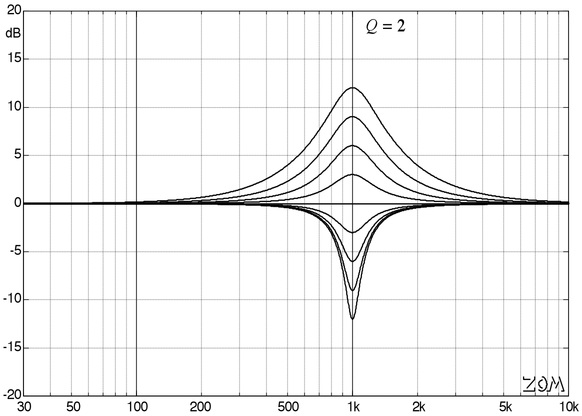


Fig. 10.3.18: Characteristic of a Constant-Q-Equalizer. The specified Q is the denominator-Q.

The constant-Q-equalizer is held in high esteem because the Q-factor does not increase as the boost-factor grows but remains constant independent of the boost. It should be added that it is the denominator-Q-factor that remains constant because the numerator-Q-factor of course does change. It is not entirely far-fetched to give priority to the denominator-Q over the numerator-Q because the **decay-coefficient** determining the time-envelope of a step- or an impulse-response indeed does depend only on the denominator-Q. However, whether it is in fact desirable that abutting EQ-bands show a boost-dependent, more or less pronounced overlap as depicted in Fig. 10.3.18, needs to be determined on a case-by-case basis according to individual preferences.

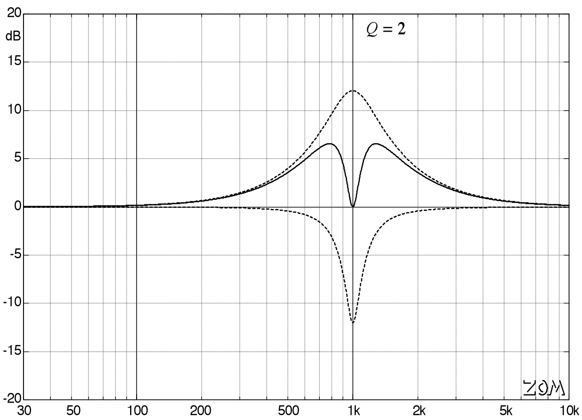


Fig. 10.3.19: Series-connection of two constant-Q-equalizers. Single filter (----) and series connection (——). For the gain to add up to 0, both Q-factors need to be reciprocal (right-hand picture).

Fig. 10.3.20 shows a circuit often utilized for designing graphic EQs. The frequency-dependent impedance *Z* of the resonant circuit may be realized in a passive (RLC) or an active manner; the latter via adding an additional amplifier. The boost-factor can be controlled with the potentiometer *P*, the center-frequency and the Q-factor are pre-set by the circuit design.

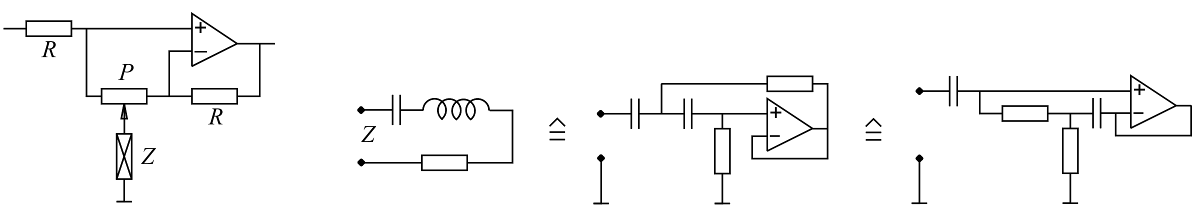


Fig. 10.3.20: Active EQ-circuit. The series-resonance-circuit (*Z*) may be realized via either active circuit. The active resonant circuits are approximations of an ideal series-resonance circuit

The circuit presented in Fig. 10.3.20 offers the possibility to vary Q (within certain limits) depending on the boost-factor (**Fig. 10.3.21**). As can be seen, we obtain inverse behavior with a bandwidth varying in detail. Relatively high impedance in the potentiometers results in the characteristic as show on the right, and low-impedance pots give the curves on the left. For linear potentiometers, the boost-value changes predominantly towards the end to the control path – therefore special pots with an S-shaped characteristic are required.

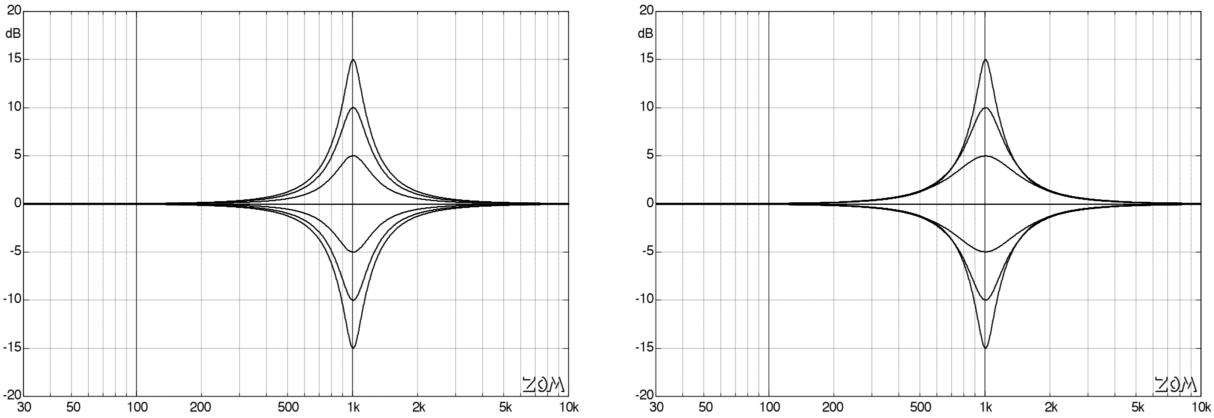


Fig. 10.3.21: Transmission characteristics of the EQ-circuit according to Fig. 10.3.20.

A multi-band graphic EQ may be designed with little effort by adding into the circuit according to Fig. 10.3.20 further potentiometers with corresponding different resonant circuits. **Fig. 10.3.22** has the corresponding diagrams for various settings.

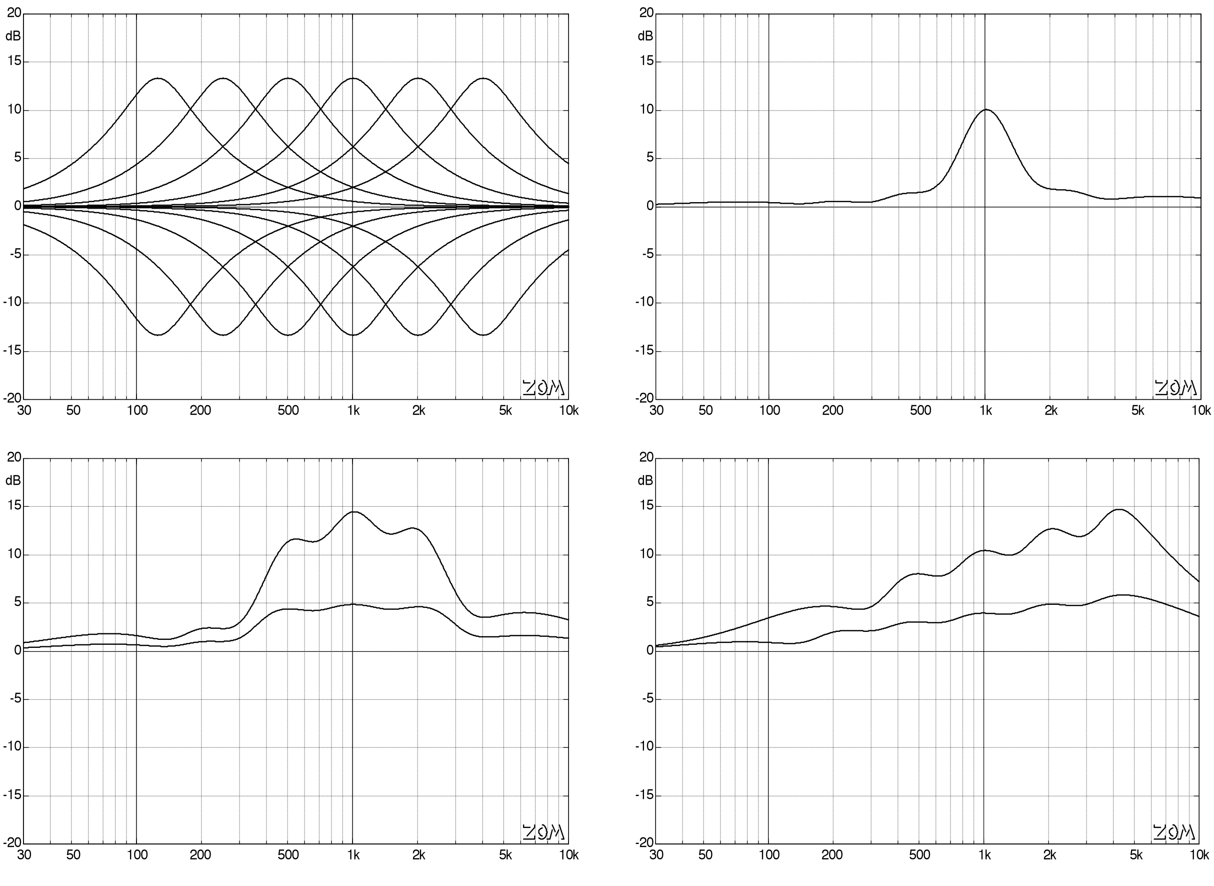


Fig. 10.3.22: Octave-equalizer: single filter (upper left). Six-band EQ, boost only in the 1-kHz-channel (u. right). Boost only in 3 bands (lower left). Boost increasing with frequency (l. right).

10.3.3 Presence-Control

In studio-electronics, the term “presence” often characterizes the frequency-range between about 1 kHz and 4 kHz, and a “**presence filter**” designates an equalizer operating in this range. In guitar amplifiers, however, the presence-control represents an alternative to the treble-control. An early variant of the presence-control is found in Leo Fender’s Bassman: already the early versions (e.g. 5B6) include negative feedback in the power amplifier, and this becomes frequency-dependent in the model 5D6. Presumably an additional treble boost was desirable. There already was a treble-control so a different designation had to be found: presence-control.

Having picked the Bassman as a model for his JTM-45, Jim Marshall (or rather Jim’s tech Ken Bran) adopts this presence-filter, as well. Only VOX takes the opposite approach: since the AC-30 already boosts the treble almost too much, the power amp here receives a treble-attenuator designated with “Cut”. In the Fender- and Marshall-amps, the presence-filter operates on the basis of a simple **principle**: a low-pass integrated into the negative-feedback-loop diminishes the loop-gain for high frequencies, and boosts the treble that way. However, despite their simple function, the circuit includes two special aspects. First, the **loudspeaker** needs to be considered as part of the negative-feedback-loop: its impedance contributes to the effect of the presence-filter. Second, the power-amplifier of a guitar amp is often subject to **overdrive**. The presence filter becomes part of a non-linear system the tonal effects of which are different from those of the treble-control.

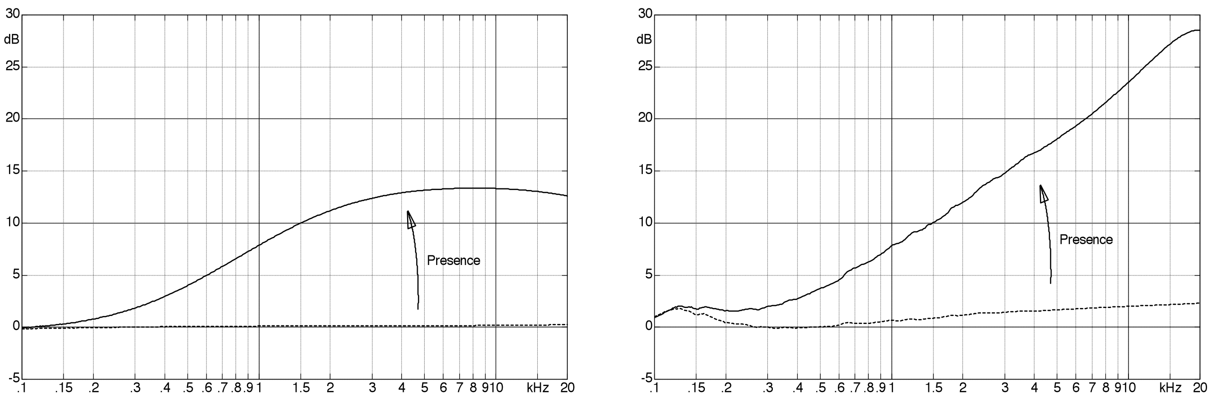


Fig. 10.3.23: Effect of the presence filter in the Marshall JTM-45. In the measurement on the left, the 16-Ω-output was loaded with a 16-Ω-resistor whereas on the right the load was a 4x12 speaker box (1960 AX).

In **Fig. 10.3.23** we see measurements on the JTM-45. The generator-signal was fed to the input of the differential amplifier; measurements were taken at the output of the power-stage. In one case the load was a 16-Ω-resistor; in the other case a loudspeaker-box was used. The latter is specified at 16 Ω, as well, but does not have constant impedance; rather, its impedance is frequency dependent.

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**).).

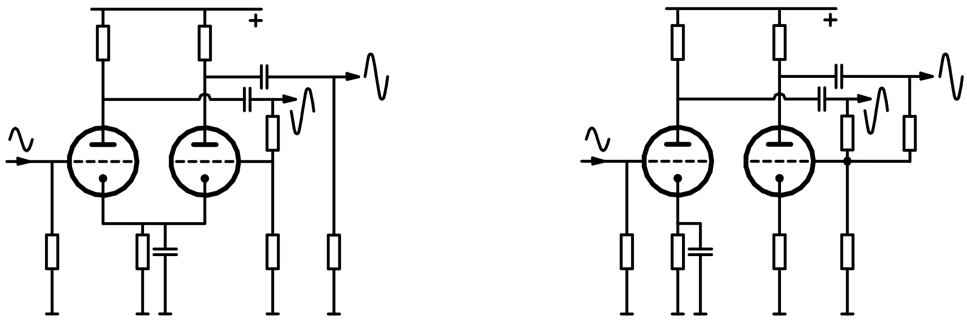


Fig. 10.4.1: Phase-inverter in common-cathode configuration. Right: modified version with negative feedback.

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).

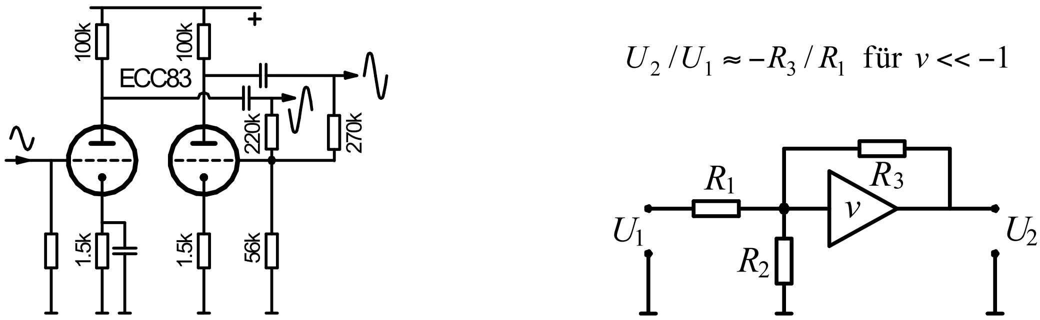


Fig. 10.4.2: Paraphase-circuit with current/voltage negative-feedback (Fender Deluxe 5D3, 1954).

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.

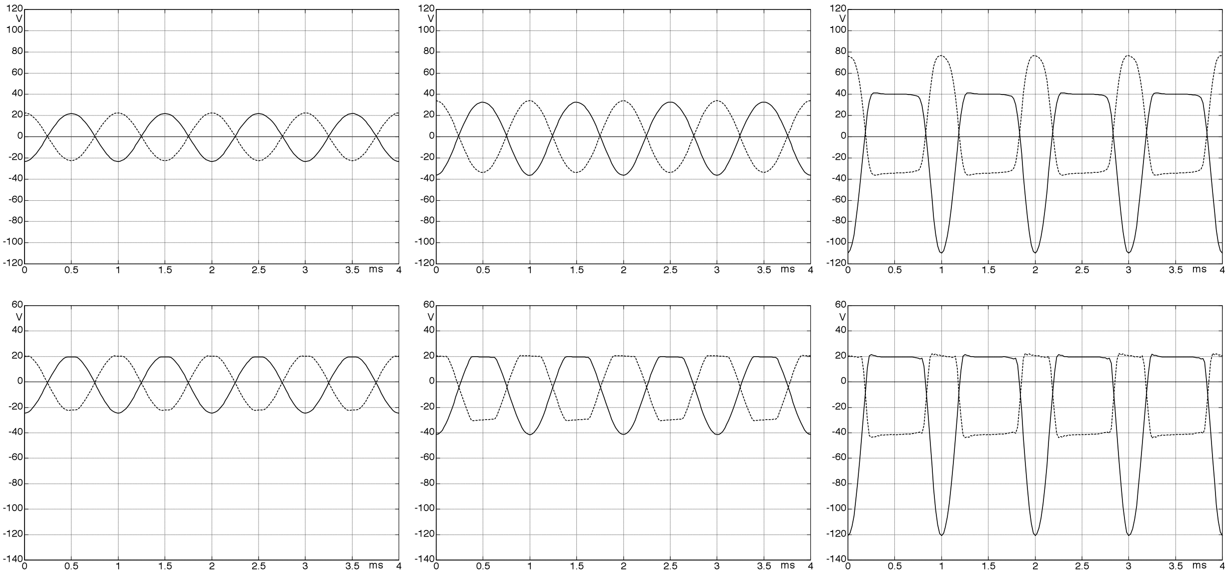


Fig. 10.4.3: Measurements on a paraphase-stage **<u>without</u>** negative feedback: 1st tube (——), 2nd tube (---).
Top: no grid-current limiting. Bottom; grid-current happening from 20 V. Supply-voltage for the triodes: 260 V.

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.

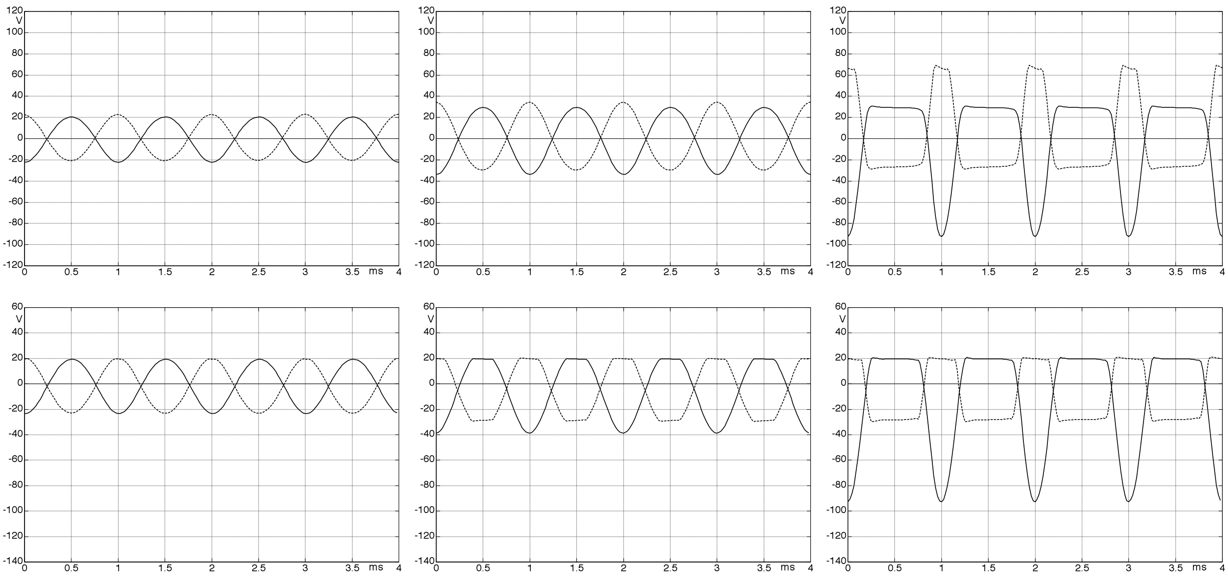
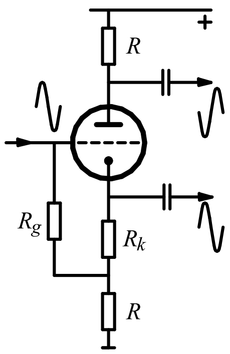


Fig. 10.4.4: Measurements on a paraphase stage **<u>with</u>** negative feedback: 1st tube (——), 2nd Tube (---).
Top: no grid-current limiting. Bottom; grid-current happening from 20 V. Supply-voltage for the triodes: 235 V.

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.



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

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

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

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.

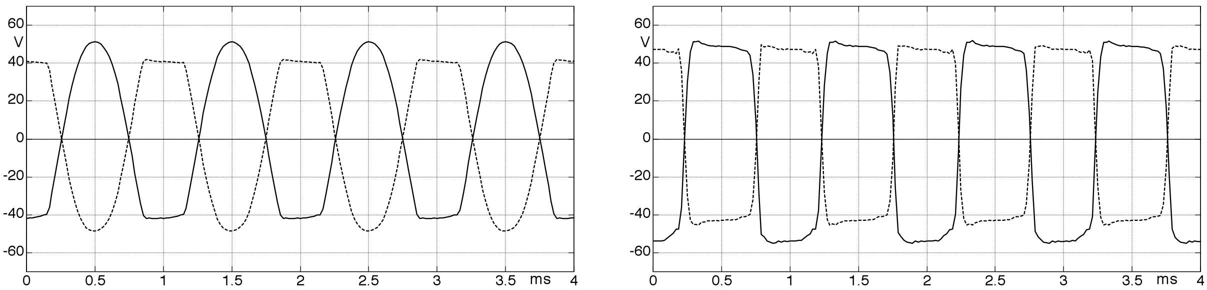


Fig. 10.4.6: Cathodyne-stage without load; AC-component. Plate-voltage (----), cathode-voltage (——).

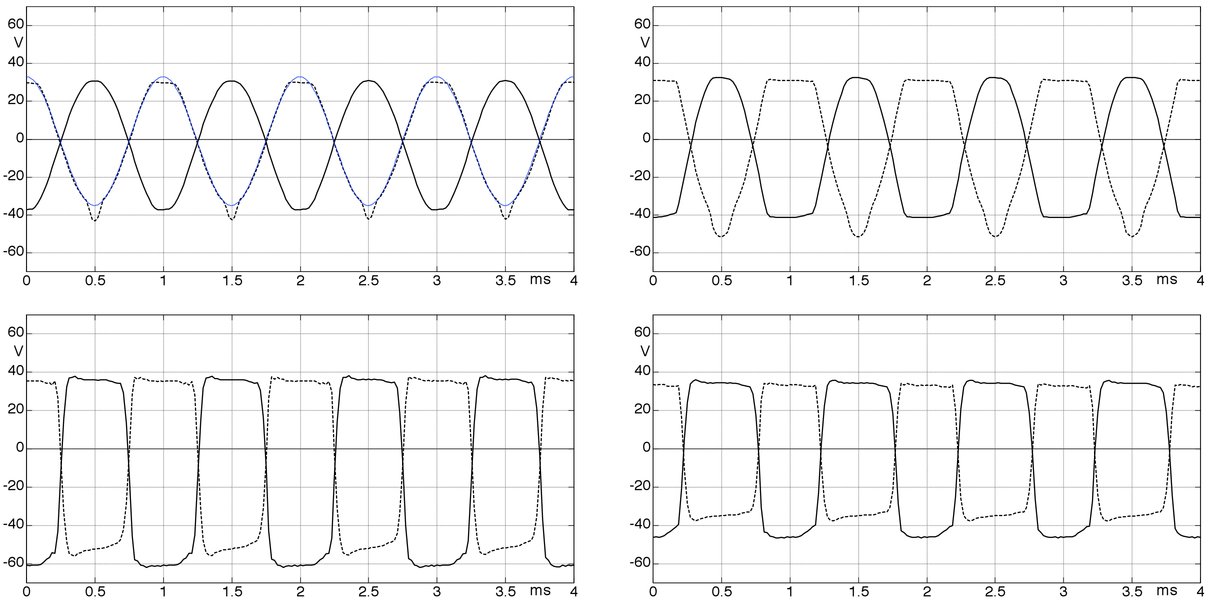


Fig. 10.4.7: Cathodyne-stage with load; AC-component. Plate-voltage (----), cathode-voltage (——).
The bottom right-hand picture shows the situation after longer-term overdrive.

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.

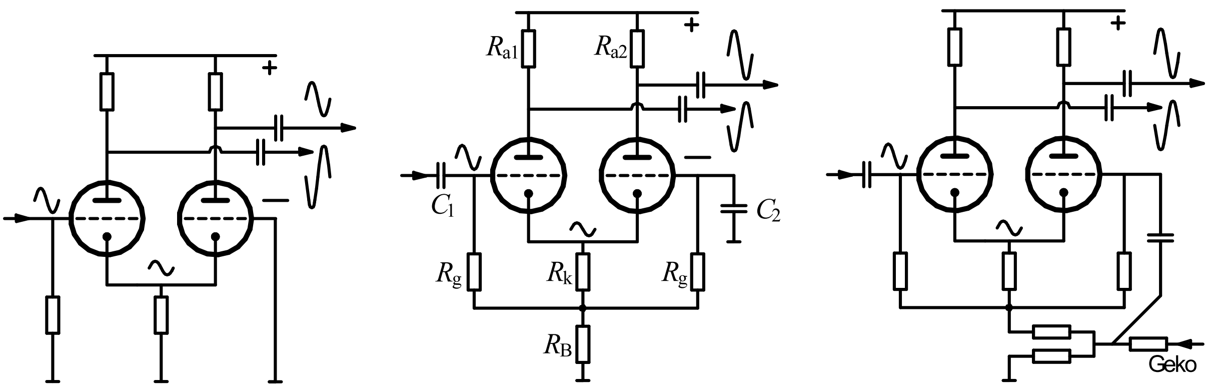


Fig. 10.4.8: Differential amplifier with negative feedback via the cathode (Geko = negative feedback).

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.

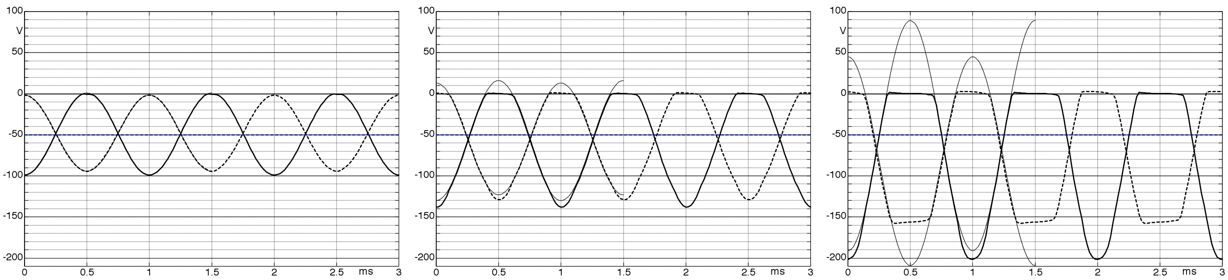


Fig. 10.4.9: Measurements at the differential amp of a Fender Super-Reverb (AB-763, negative feedback deactivated). Power-tube bias = –50V. Grid-voltage of the 1st power tube (V7 = ——), and of the 2nd power tube (V8 = ----). On the left, undistorted cosine-oscillations are shown for comparison.

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.

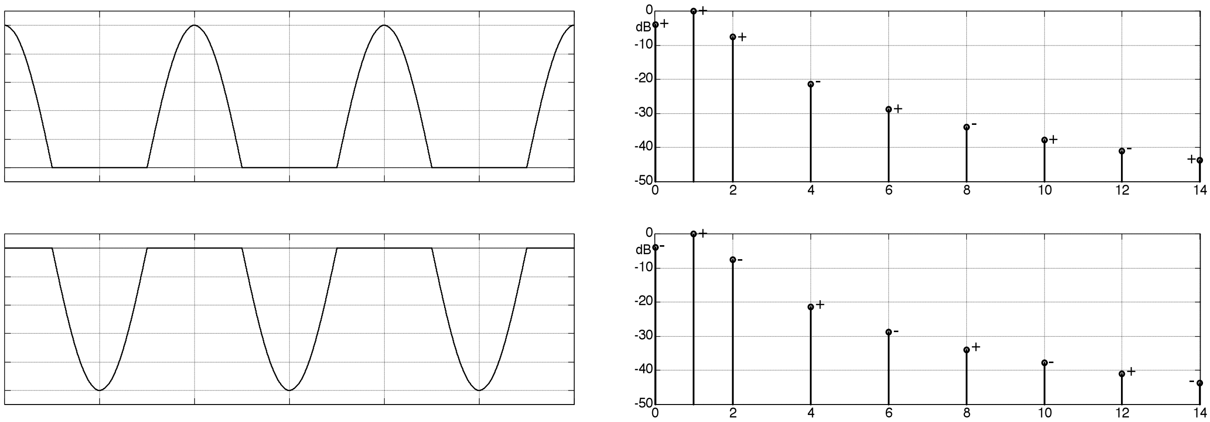


Fig. 10.4.10: Time functions (left) and spectra of the half-wave signals. The signs of the Fourier-components are the same only for the 1st harmonic, and consequently only this component remains after the addition.

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).

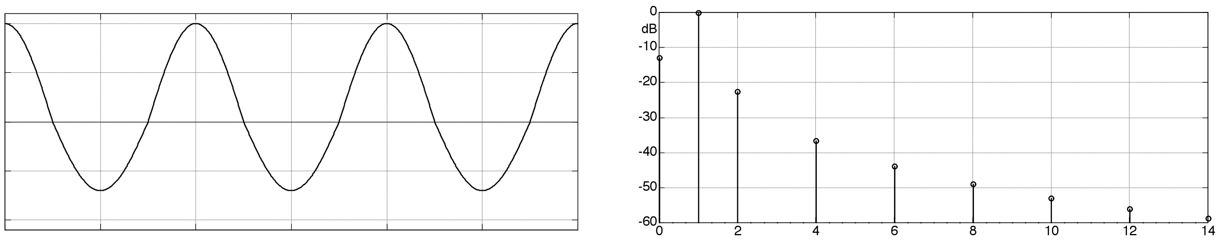


Fig. 10.4.11: Time function and spectrum of a signal with different amplification of the two half-waves.

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.

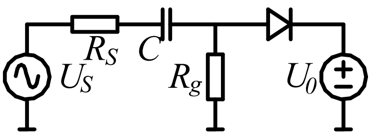


Fig. 10.4.12: Simple model-circuit to simulate grid-currents.

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.

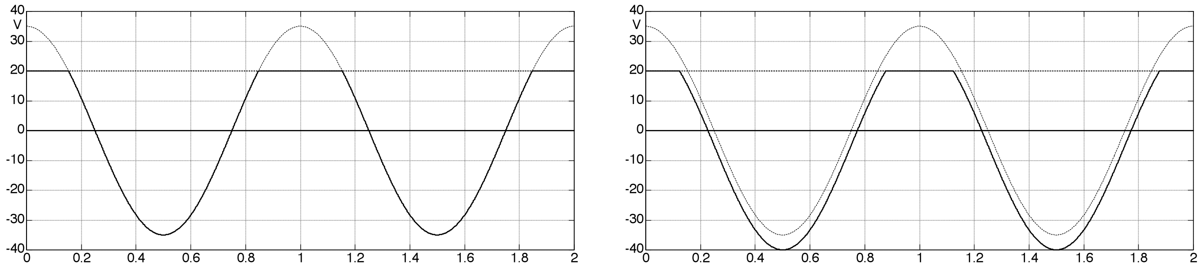


Fig. 10.4.13: Potential-shift due to grid-current in the output tubes. Left: AC-voltage limited to merely 20 V; right: AC-voltage limited and shifted (capacitor-polarization).

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.

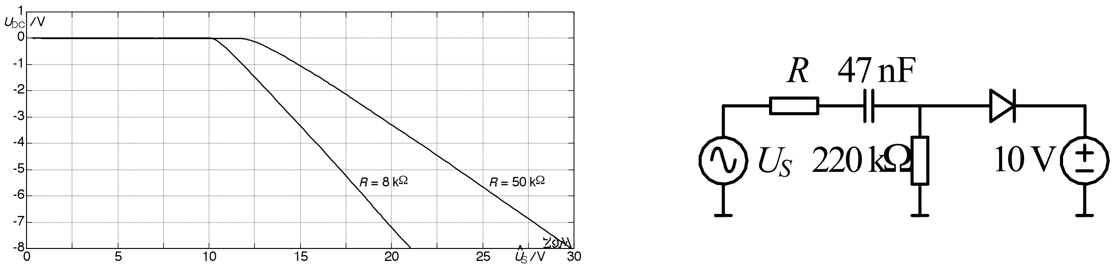


Fig. 10.4.14: Average grid-voltage-bias U_{DC} in dependence on the drive-voltage-amplitude (model).

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.

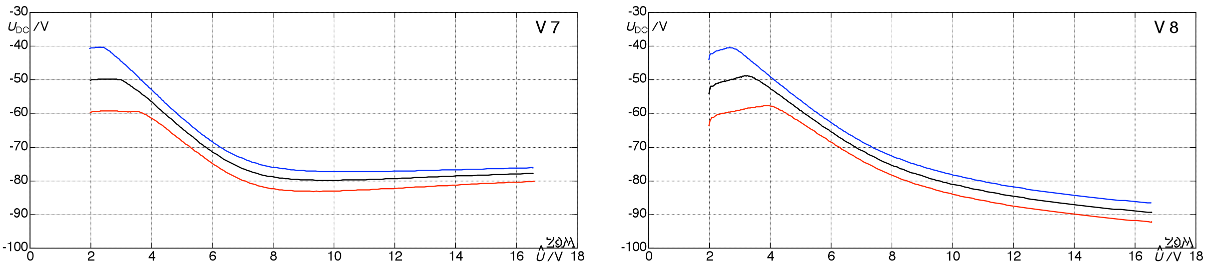


Fig. 10.4.15: Fender Super-Reverb, grid-bias-voltage of output tubes (mean); 3 different operating points. Drive voltage (abscissa) is the grid-voltage of the left-hand differential-amplifier tube.

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 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.

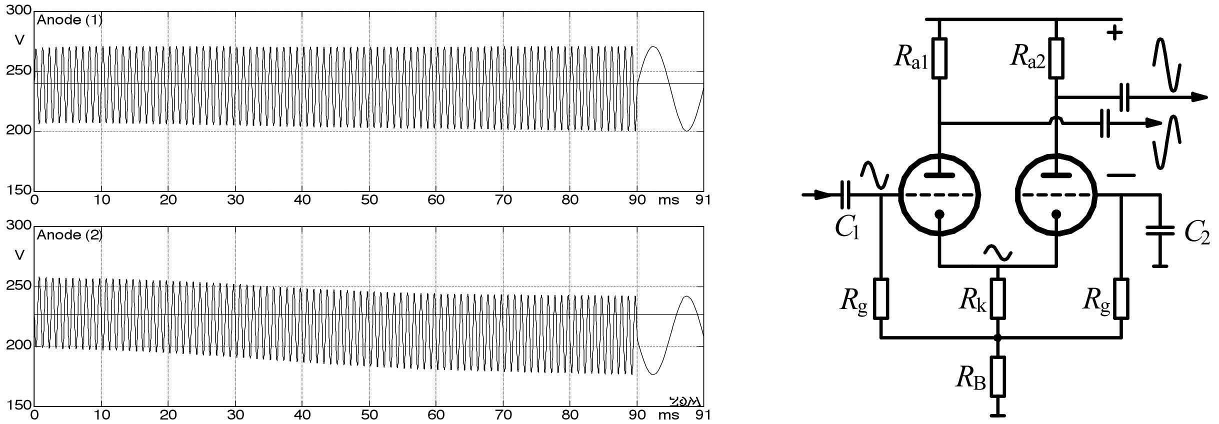


Fig. 10.4.16: Shift of the operating point in the differential amplifier of a Super-Reverb (negative feedback deactivated). The mean-value of the plate-voltage for the right-hand triode shifts towards lower voltages.

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.

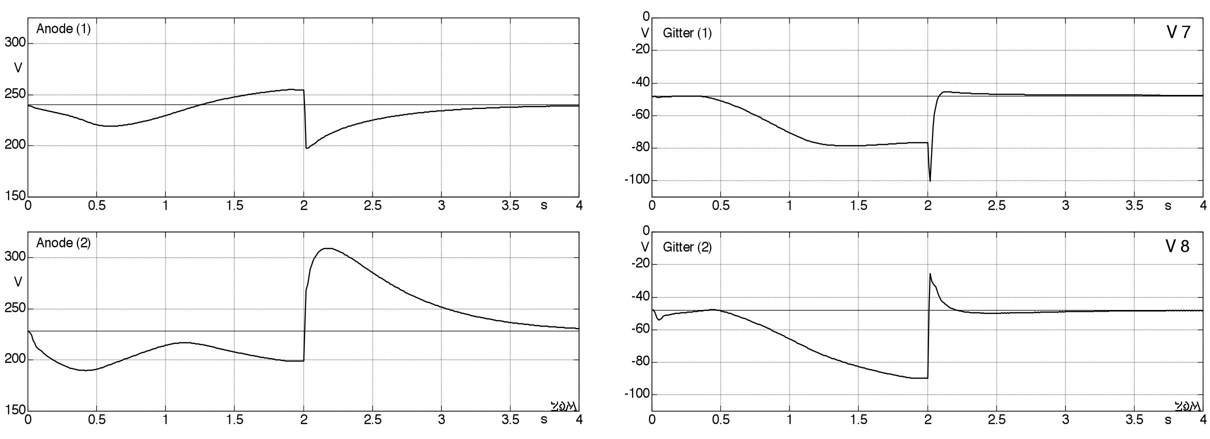


Fig. 10.4.17: Mean values of the voltages at the plates of the differential amplifier (left) and at the output tube grids. During $0 < t < 2$ s, the signal level rises by 20 dB, at t = 2 s the signal is shut off. Super-Reverb.

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.

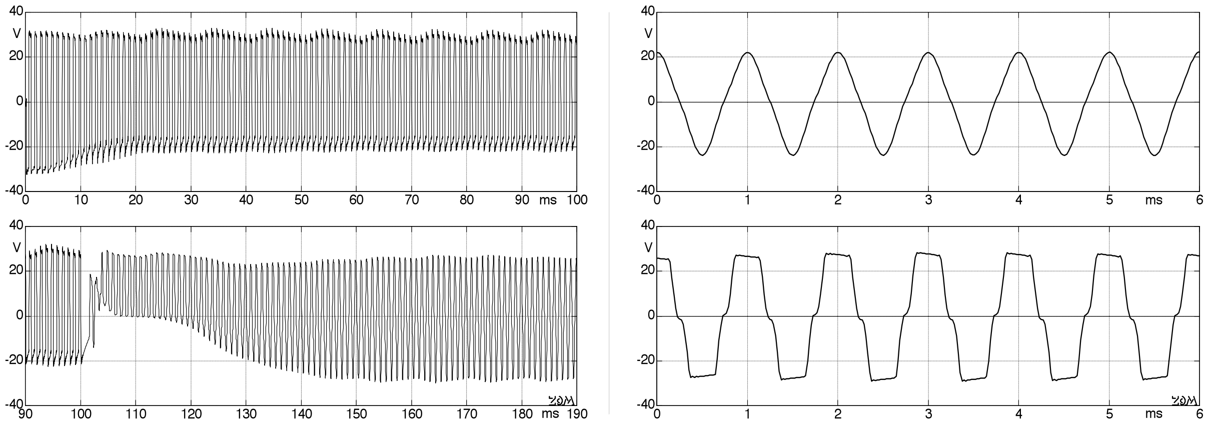


Fig. 10.4.18: Super-Reverb, loudspeaker-voltage (overall feedback-loop deactivated).

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.

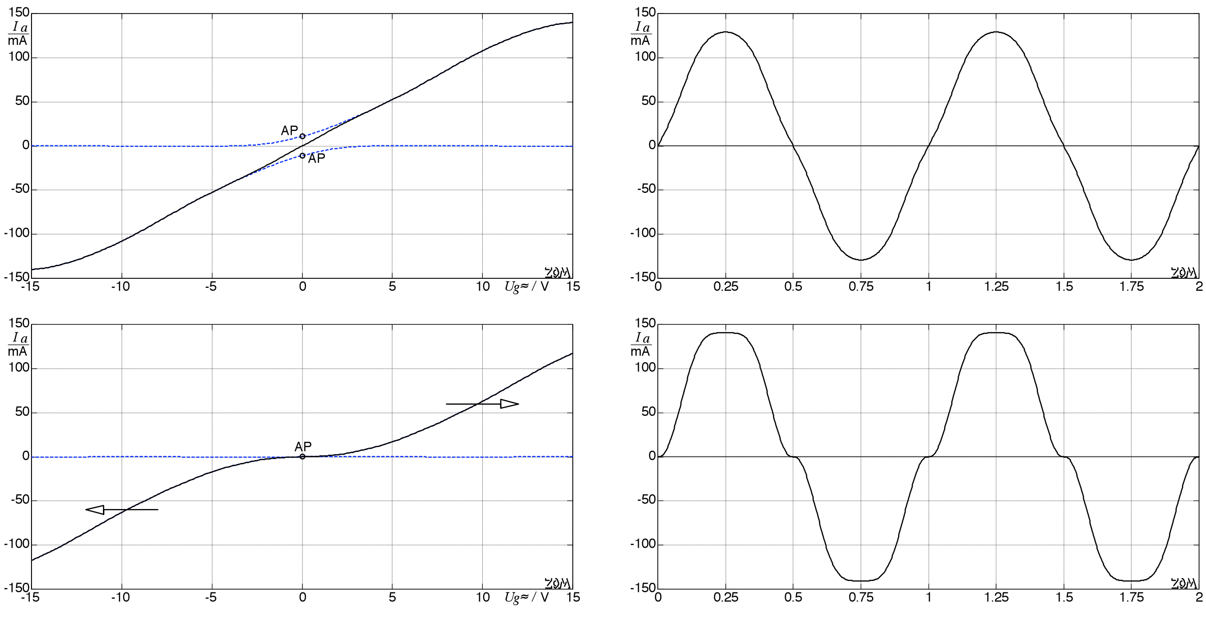


Fig. 10.4.19: Dynamic (drive-level dependent) crossover distortion (compare to Chapter 10.5.8).

♣ The power-amplifier still remains overdriven