10.5 Power-Stage

The power-stage is the last amplification stage in the signal chain; it delivers the power required to drive the loudspeaker. In most cases, it operates with rather pathetic efficiency because normally less than half of the power produced by the power stage is actually fed to the loudspeaker – the remainder is converted into heat within the power-tube(s). In order to be able to sufficiently dissipate this power loss, the tube(s) deployed in the power-stage is (are) larger than typical preamplifier tubes. The thermal load capacity of power tubes typically amounts to 12 – 45 W with their physical volume up to 10 times that of a preamp-tube. Since power-tubes can deal with high voltages but not with high currents, they are almost never directly connected to the loudspeaker. Rather, the plate-currents of the power-tubes are fed to the output transformer that takes care of an impedance matching towards the speaker.

A good overview is provided by the **family of output characteristics** of the power-tube (**Fig. 10.5.1**) showing the relation between plate-voltage and plate-current. Multiplying these two quantities yields the **power-dissipation at the plate** P_a, i.e. the power heating up the plate of the tube (in addition to the heating done by the tube filament). If the specified maximum dissipation at the plate is exceeded for long periods of time, the tube begins to glow and may be destroyed. The so-called **power-hyperbola** is given in Fig. 10.5.1 as the dashed line, indicating the largest permissible plate-current for the respective plate-voltage. To the right, the characteristic finds it limitation in the largest allowable plate-voltage; larger values will cause sparking and damage. Towards the top, the maximum specifications of plate-current and/or grid-drive provide a ceiling; the lower limits are given by the blocking behavior of the tube. Normally, tubes are rather good-natured regarding overload situations (much more so than transistors) because the associated thermal time constants are much longer. However, this behavior must not be interpreted as general "indolence": continuous overload will reduce the lifetime (Chapter 10.5.9).

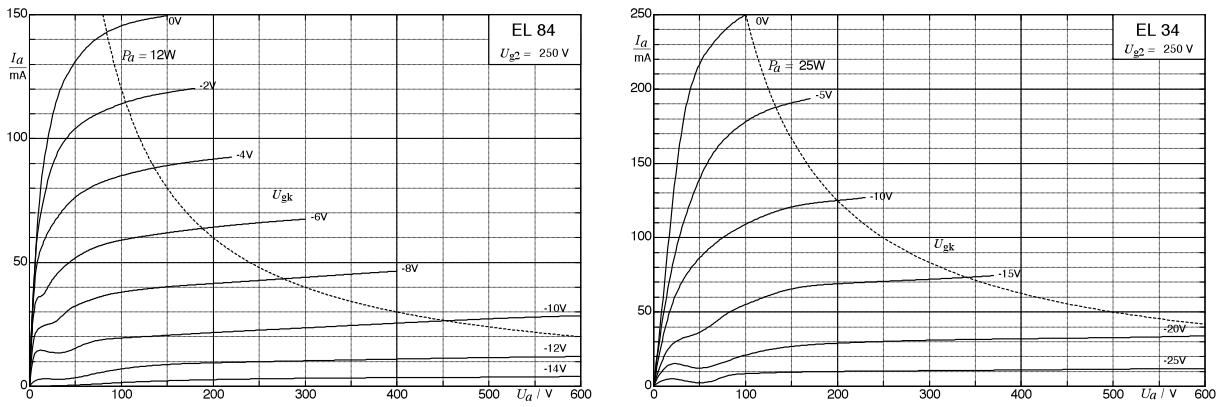


Fig. 10.5.1: Family of output characteristics of two typical power-pentodes. Screen-grid-voltage U_{g2} = 250V.

It should be noted that the characteristics given in Fig. 10.5.1 are sourced from datasheets (as is the case for all tube characteristics); to a degree, the individual tube-specimen will look different. In addition, it needs to be considered that power-tubes are almost always tetrodes or pentodes, and consequently their behavior is defined by both control-grid *and* screen grid. For data-sheet specifications, the screen-grid-voltage is assumed to be constant – however, reality shows that it depends on the drive-levels, after all. On the one hand, this is due to the fact that the supply-voltage drops somewhat as the drive-levels increase ("sagging"), and on the other hand, it is because there is a voltage-drop across the grid-resistor.

10.5.1 Single-ended (class A)-operation, tetrode, pentode

In the single-ended, class-A power-stage, one (single) power-tube operates in common-cathode configuration with the output transformer being part of the plate circuit (transformer-coupling). Without AC-drive ("quiescent state"), a stable balance appears – it is called the operating point (**OPP**). The characteristics shown in **Fig. 10.5.2** yield an OPP at 250 V and 48 mA, if a voltage of -7.5 V between (control) grid (g_1) and cathode is chosen. This can be done e.g. by using a cathode-resistor of 142 Ω. The cathode-current (the sum of the 48-mA-plate-current and the 5-mA-screen-grid-current) will then generate a positive cathode-voltage of + 7.5 V (relative to ground). With the control-grid at ground-potential (U_{g1} = 0) a control-grid-to-cathode-voltage of -7.5 V results (i.e. the control grid is negative vs. the cathode).

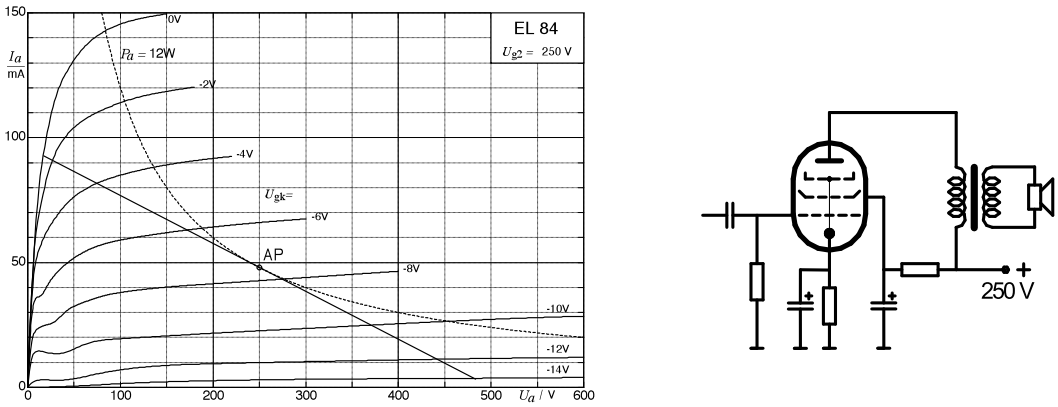


Fig. 10.5.2: Output characteristics of the EL84, power-stage circuit (single-ended class-A operation). AP = OPP

As a drive signal appears ($U_{g1} \neq 0$), plate-voltage and –current change. As a first approach, it will be sufficient to consider the transformer in the plate-circuit as a large inductance connected in parallel with an ohmic resistor (Chapter 10.6). In this model we have only pure DC flowing through the inductance, and only pure AC flowing through the resistor. With a drive-signal present, the U_a/I_a-point will move along the **load-line** given in Fig. 10.5.2: as the grid-voltage is enlarged, the plate-current increases and the plate-voltage drops until a limit is reached at 17 V / 92 mA with U_{gk} = 0. **Fig. 10.5.3** shows that the relation between input- and output-magnitudes is non-linear: merely with small drive-signals around the operating point we can obtain an approximate image of the input signal with small harmonic distortion. In addition, we need to bear in mind that in reality, the power-tube is rarely driven via a low-impedance source. Often, the driver-tube ahead of the power-tube is operating in common-cathode configuration i.e. with a relatively high internal impedance (e.g. 50 kΩ) – in this case the grid-current of the power-tube already distorts the control (drive) signal.

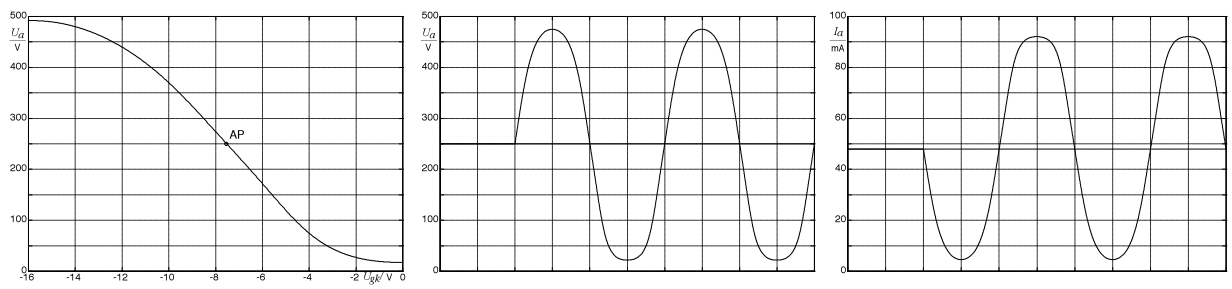


Fig. 10.5.3: Transmission characteristic; plate-voltage and plate-current for sinusoidal U_{gk} (from a stiff voltage source).

The output transformer takes the AC-component from the plate-circuit and generates the loudspeaker-current that is enlarged by the turns-ratio factor *TR* (secondary current, Chapter 10.6). The AC-load of the power-tube results from the inclination of the load line; in Fig. 10.5.2 this is 5208 Ω (19.2 mA / 100 V). From this load-impedance at the plate, and from the loudspeaker impedance (e.g. 8 Ω), we get a first approximation of the transformation ratio (ratio of turns in the two transformer windings) *TR* of the transformer: $TR = \sqrt{5208 / 8} = 25.5$. In view of the transformer losses, this value should be decreased by about 10 % – now we arrive at approximately $TR \approx 23$ [for more exact calculations see e.g. Schröder, Vol. II].

In quiescent state (i.e. without any drive signal), the plate is at 250 V and the plate-current is 48 mA. Multiplying these two values gives us the dissipation loss at the plate of P_a = 12 W. Since the (idealized) load-impedance was assumed to be an R/L parallel-circuit (= short-circuit for DC), the supply voltage is calculated as U_B = plate-voltage + cathode-voltage = 257.5 V. This is a bit too much "lab jargon" and we need to get more precise. What the data books term "plate-voltage" is in fact the voltage drop U_{ak} between plate and cathode; it is also called plate/cathode-voltage. In a series connection to it we have the voltage drop occurring across the cathode resistor, also termed cathode-voltage: $U_B = U_k + U_{ak}$. Without drive signal, the cathode resistor (142 Ω) absorbs 0,4 W while the plate absorbs 12 W, and the screen grid 250 V · 5 mA = 1.25 W. Consequently, the power supply needs to deliver, in **quiescent state**, 13.65 W. **With a drive signal**, the plate-current becomes time-variant und oscillates between two limit-values, e.g. 5 und 92 mA (Fig. 10.5.3). If we ignore the non-linear re-shaping, the average of the current remains constant, which implies: the power that the power supply needs to make available is approximately constant i.e. independent of the drive signal level! Multiplying the AC-components of the plate-voltage and the plate-current (Fig. 10.5.3) results in the **effective power** pushed into the load-impedance: P_N = 6 W. Given an ideal transformer, this power fully arrives at the load-impedance (the loudspeaker); in reality a loss of 20% is likely. Only about 4.8 W arrive at the loudspeaker and the remaining 1.2 W are converted into heat in the transformer.

In summary: the power supply needs to deliver about 14 W independently of the drive signal, which leaves just under 5 W output power at full drive level – with the output signal being already subjected to substantial non-linear distortion (strong THD). The efficiency of this circuit is 35%, at best – or even as low as 26% if we include in our considerations the tube heating. The latter is necessary to operate the EL84, and gobbles up another 4.8 W.

As inefficient this circuit may be – it was indeed used in some early guitar amplifiers. One of the first VOX-amplifiers, the AC-4, generated 4 W from a single EL84 in a single-ended class-A configuration. The first smaller Fender amps uses the single-ended Class-A circuit, as well – we find it e.g. in the Champ, Bronco, Princeton and Harvard amps, although these used the 6V6-GT, a 12W beam-power tetrode rather than the EL84. Over the years, the Fender amps in particular were subject to various modifications. Among these the increase in **supply voltage** is especially striking: early versions had 305 V; an increase to 305 V followed, and finally there was as much as 420 V. Can we boost the output power that way? Which is the optimum operation point to achieve the maximum power output? Which load impedance is optimal for the tube? Using simplifications in the tube- and transformer-data, the calculation for optimum working conditions is unproblematic. However, in the real world one needs to consider deviations from these ideal conditions. In particular the maximum current-load of the power tubes is subject to manufacturing tolerances, and transformer losses (build-size!) determine the eventually achievable output power, too.

With the idealized assumption that, in the plate-circuit of the power tube, the power-hyperbola is the only limitation, the left section of **Fig. 10.5.4** shows two load-lines that each are tangents to the hyperbola. The division U_{AP} / I_{AP} yields the optimum operating point (OPP), corresponding at the same time to the negative slope of the hyperbola at the OPP. The maximum possible voltage deflection at OPP_1 is 400 V_{ss}, resulting in 6 W, with a load resistance of 3333 Ω. The same power can be achieved in OPP_2: the voltage deflection is indeed larger at 600 V_{ss}, but the current is correspondingly smaller. If we define the power hyperbola as limit, the achievable maximum power is always exactly half of the maximum dissipation-power at the plate – independent of the OPP. For a **real circuit** we need to factor in that the plate-current cannot become indefinitely large. In the right-hand section of the figure, the output characteristic of a 6V6-GT is indicated as limit for the case that the grid/cathode-voltage is zero. This curve must not be seen as the absolute limit – even larger plate-currents would be possible if the grid/cathode-voltage were positive. However, the typically used drive-circuits could not deliver the necessary current, and consequently it is purposeful to define, in addition to the power hyperbola, $U_{gk} = 0$ as the limiting factor. Now, the maximum voltage deflection reachable at OPP_1 is not 400 V_{ss} anymore but decreases to 334 V_{ss}, and the OPP is not located in the middle of the load line any longer. A conducive shift of OPP_1 from 200 V to 233 V does enable us to establish symmetry with regard to the maximum drive level. However, the reduction of the maximum voltage deflection by 16.5% decreases the maximum power-offering by 30% (in our example from 6 W to 4.2 W). For OPP_2, the reduction of the voltage-deflection makes itself less strongly felt (5.6 W instead of 6 W), and we can expect the operation with a higher voltage to bring somewhat **more power**.

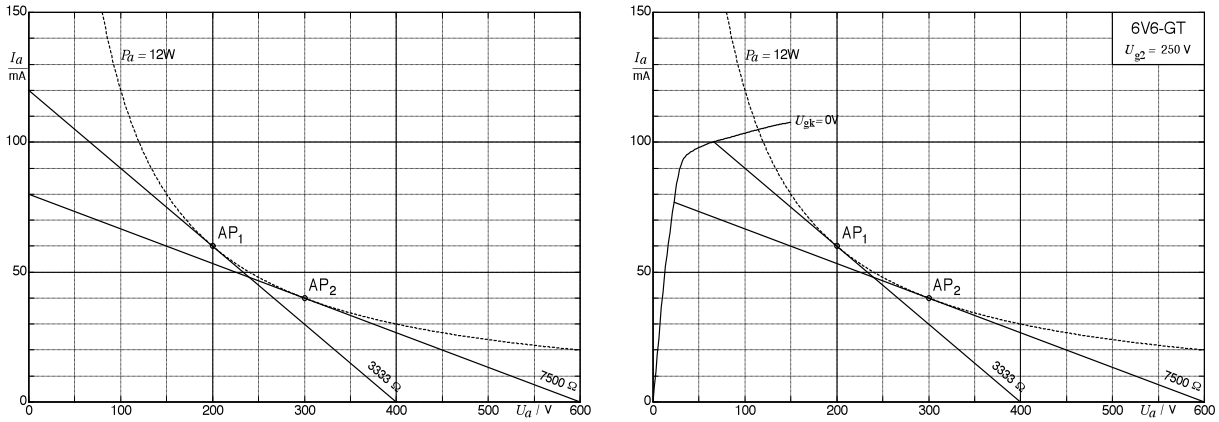


Fig. 10.5.4: Output characteristic with two different operating points; the power hyperbola is the limit.

The above calculations regarding the achievable power-output deliberately are of a rather "principle" character in order to illustrate basic functions within the power stage. If we do not consider the power hyperbola as limiting factor, the circuit will deliver 50% of the maximum power dissipated at the plate to the output transformer – irrespective of the tube used. This upper power-threshold can only become smaller (and never bigger) as individual tube-limit-data are incorporated. Besides the maximum power-dissipation at the plate, in particular the maximum tolerable plate-voltage and the maximum allowable power-dissipation of the screen grid need to be considered. For a supply voltage of 300 V, up to 600 V may occur at the plate, and even as much as 840 V for 420 V supply voltage. Also, even higher voltages may appear, since the load impedance (loudspeaker) is not a purely ohmic 8-Ω-resistor but will become inductive (and thus larger) at high frequencies. Even if the insulation within the transformer is exemplarily well done: at too large voltages, arc-over is possible in or at the tube, and it can lead to destruction.

So much for an introductory, basic description of the behavior of a single-ended power-stage – now on to the details. For the **triodes** deployed in preamplifiers, a simple power law was formulated as an approximation (Child/Langmuir, Chapter 10.1.3):

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

The plate-current I_a depends on the grid/cathode-voltage U_{gk}, on the plate-voltage U_a, and on the open-loop gain μ, the latter also known as **durchgriff**: $D = 1/\mu$. To get even more into detail: the free conducting electrons in the metal cathode are highly mobile but cannot leave the metal in its cold state. A special coating combined with red-heating enables a significant portion of the electrons to leave the metal and form, in the immediate vicinity around the (heated) cathode, a kind of "electron-mist" – also called "space-charge cloud". The more electrons accumulate in front of the cathode, the more negative this **space-charge area** becomes, and the more effectively further electrons are inhibited to move against this negative potential – an equilibrium results. A positively charged plate will superimpose an electron-accelerating plate-field over the electron-inhibiting space-charge field, and the former field will suck electrons away from the cathode and draw them to the plate. The space-charge decreases, enabling more electrons to leave the cathode. The electrons leaving the cathode form the cathode-current, and the electrons arriving at the plate form the plate-current. A **(control-) grid** (three-electrode-tube = triode) introduced between cathode and plate will create, via its electrical potential, an additional field. Consequently, on top of the space-charge field, *two* fields that are controllable via the electrodes act on the electrons and therefore influence the current: one generated by the (control-) grid, and the other generated by the plate. Since the grid is positioned closer to the cathode, it exerts the bigger influence: the plate needs to first "reach through the control grid to the space-charge" – hence the term "durchgriff" (the term taken from German, meaning "reaching through"). For the ECC83, the datasheet indicates a rather small value at $D = 0.01$. However, with the plate-voltage being about 100 times the value of the grid/cathode voltage, both U_a and U_{gk} influence the plate-current. Textbooks on practical circuit-design see the grid as control-electrode and designate U_{gk} as control-voltage. More theoretically oriented oeuvres combine the summands $U_{gk} + D \cdot U_a$, using the same term **control-voltage** for the combination i.e. this term may have two different meanings! In the formula above, U_{St} is the theoretical control-voltage considering both the influence of grid *and* plate, with K_2 being a tube-specific constant.

One may consider it a problem that the plate-current of the triode does not only depend on the grid/cathode-voltage but also on the plate-voltage. A solution can be found by inserting an additional **screen grid** (g_2) between control grid (g_1) and plate, and connecting it to a high positive voltage – this way the electrons are predominantly accelerated by the control grid and the screen grid, with the plate-potential retaining merely a minor significance. For the resulting 4-electrode tube (= **tetrode**), the potentials of all electrodes can be described via a theoretical control-voltage:

$$U_{St} = U_{g1} + D_1 \cdot U_{g2} + D_1 \cdot D_2 \cdot U_a \qquad \text{Control-voltage of the tetrode}$$

The tube parameters D_1 and D_2 – both considerably smaller than 1 – can again be interpreted as durchgriff. $D_1 \cdot D_2$ shows the (intended) small influence of the plate-voltage.

As an example: if the control-grid-voltage has to change by 1 V in order to change the plate-current by 10 mA, then for the same plate-current change the screen-grid-voltage would have to be changed by 20 V, or the plate-voltage by 400 V. To map the control-voltage onto the plate-current, we could use the power law for the tetrode, as well, but we would need to introduce considerable **corrections** to obtain a good match to the actual behavior. A main reason for this discrepancy between simple theory and practice is the release of **secondary electrons** from the sheet-metal of the plate. As soon as the electrons arriving from the cathode are accelerated with more than 10 V difference in potential, they have enough energy to knock, as they hit the metal, further electrons from the plate – these are the secondary electrons. With the screen-grid-potential lower than the plate-potential, this process is not disruptive because the secondary electrons return to the plate. However, for higher screen-grid-potential the secondary electrons fly on to the screen grid – correspondingly decreasing the plate-current and increasing the screen-grid-current. This is the reason why an enormous bump appears in the I_a/U_a-characteristic of the tetrode for small plate-currents. This bump is undesirable (**Fig. 10.5.5**).

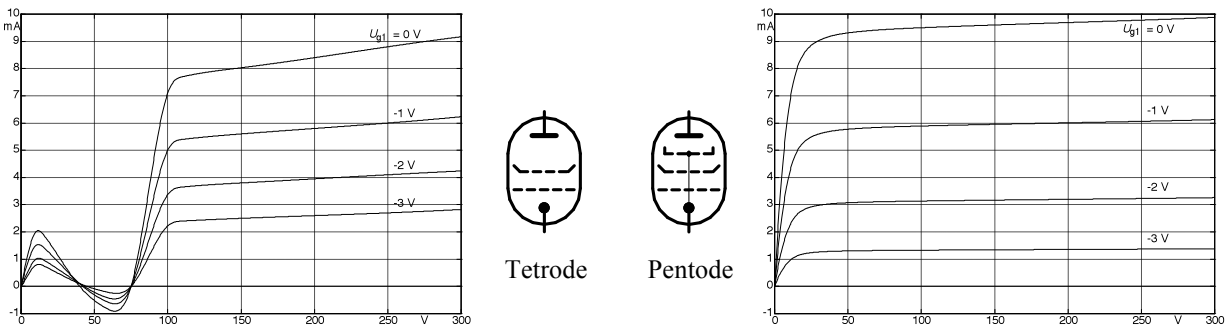


Fig. 10.5.5: Output characteristics (I_a vs. U_a) of a tetrode (left) and a pentode (right).

Corrective action is provided by yet another electrode, the **suppressor grid** (or retarding grid) located between screen grid and plate. Its job is to push back the secondary electrons en route from the plate so that they will not land on the screen grid. This only works if the suppressor-grid potential is much lower than the screen-grid potential, and therefore the suppressor grid is normally connected to ground. The fast electrodes emitted by the cathode are pretty much unaffected by the this suppressor grid while the slow secondary electrons knocked out of the plate are not able to overcome the potential difference to the suppressor grid and return to the plate. Staff at Philips developed the first commercial version of this five-electrode tube (= **pentode**), with a corresponding patent filed in 1926. For a short time, pentodes are also found in pre-amplification stages of guitar amplifiers but these were soon replaced by triodes (Chapter 10.1). In contrast, we find almost exclusively power-pentodes in the power-stages, for example the EL84 (e.g. VOX), or the more powerful EL34 (e.g. Marshall).

The London-based tube manufacturer **MO-V** (MO-Valve or Marconi-Osram Valve Co Ltd.) was not allowed to manufacture pentodes due to the patent owned by Philips, and developed (around 1933) a serious alternative to the pentode: the **Beam-Power-Tetrode**. Its baffles concentrate the electron-stream such that strong space-charges strongly deemphasize the characteristic tetrode-bump. It appears, however, that there was not that much confidence in the concept at MO-V, and the corresponding rights were sold to **RCA** in the United States. RCA used them to very successfully develop the beam-power-tetrode **6L6**, and this again forced MO-V to act all the more. They introduced the **KT-66**, the "**k**ink-less **t**etrode". Both the 6L6 and the KT-66 were manufactured in many variants that can differ substantially.

The power tubes employed in guitar amplifiers may be divided into three main groups: pentodes, British beam-tetrodes and US beam-tetrodes. Among the **pentodes**, there is the EL84 for low-power applications, and the EL34 for high power. The KT-66 and the more powerful KT-88 are the **British beam-tetrodes**. Their **American counterparts** are the smaller 6V6 and the larger 6L6. All these tubes have undergone multiple redesigns since their introduction to the market; that is why we cannot speak of "the" 6L6. First came the development step from steel- to glass-container, then there were changes in the shape of the container, but also in the electrodes and thus in the electrical parameters. The RCA 6L6-GB is rated with a maximum plate-dissipation of **19 W**, the Tungsol 6L6-GB is rated at **22 W**. Can the Tungsol-tube therefore carry a higher load? That is difficult to say, because we read in the RCA datasheet: *Design-Center Values*, but in the information by Tungsol: *Design Maximum System* (more about these rating systems in Chapter 10.5.9). The Sylvania 6L6-WGA is specified at **19 W** (*Design Center*), but also at **21 W** (*Absolute Maximum*). As a first approximation, these are all tubes that are the result of a development from the 6L6, via the 6L6-G and the 6L6-GA, to the 6L6-GB, that related predominantly to the shape. Only for the **6L6-GC** we see a pronounced power upgrade to a plate dissipation of **30 W** (*Design Maximum Values*); this is probably based on changes in the metal sheet of the plate. None of these tubes were developed specifically for guitar amplifiers – that market was much too small at the time. Rather, we read: *For Radio Receivers*. There were also particularly robust *military tubes* designated with a supplementary W, e.g. the **6L6-WGB**. The corresponding electrode-build was optimized to withstand the stringent MIL-testing.

The **KT-66** is the British counterpart to the 6L6. It is specified with a maximum plate dissipation of 25 W in the Osram data-sheet; we find the same data in the Marconi data-sheet, and checking the info from MO-V yields 25 W (Design Max) or 30 W (absolute Max), respectively. **MO-V** is the moniker for the Marconi-Osram-Valve-Company, that offered the KT-66 globally under the **GEC** label. This is GEC = General Electric Corporation of England, not to be confused with General Electric USA. Both the 6L6 and the KT-66 are beam-tetrodes, i.e. tubes without a suppressor grid. Because the beam-forming sheets can be seen as a fifth electrode, after all, these tubes are often labeled as pentodes, too (despite the lack of an actual suppressor grid). The **EL34**, however, is a *true* pentode, specified at 25 W – or at 27.5 W ("at maximum drive level"). All these tubes show similar data regarding the maximum load, but we may not conclude that they can be arbitrarily interchanged – their control characteristics show considerable differences, after all.

Before we delve more deeply into the area of tube characteristics, let us take a short look at other power-tubes. Around 1950, Tung-Sol develops the **5881** and advertises it as an advancement of the 6L6 (or the 6L6-GA). In 1962, the maximum plate dissipation of the 5881 is still specified at 23 W (Design Center System) – but by that time, the 6L6 has enjoyed the further development into the 6L6-GC (30 W), and the 6L6-WGB (26 W) has been available at least since 1955. It is not surprising that not everybody regards the 5881 as the "better 6L6". And then: what does "better" mean in this context? Is this from the point-of-view of the MIG-pilot demanding full function even after a rough touchdown? Or from the point-of-view of the aficionado of classical music expecting the least possible distortion? Or from the point-of-view of the Jazz guitarist having just discovered that the tone control does not *have* to be stuck at "0" all the time? Or from the point-of-view of the Eddie-epigone overdriving his equipment (his "rig") exactly "VH-like"? To state "*the 5881 is the better 6L6*" is just as misguided as "*6L6 = KT66 = 5881*". "The" 6L6 does not exist, just as "the" KT66 or "the" 5881 do not exist. It is not just that the datasheets indicate differences – today many a KT66 internally is but a 6L6-variant.

When evaluating tubes in general, and power tubes in particular, two criteria offer themselves: the **sound**, and the operational lifespan. Sure, price and availability also figure – but we will tackle that later. The lifespan may be five hours or five years; it has its own chapter dedicated to it (Chapter 10.5.9). The sound is advertised with "powerful bass" or "clear treble", and consequently many guitarists presume that tubes would feature a frequency-dependent transmission characteristic – like that found in a loudspeaker. However, this assumption is not correct as such: tubes can process frequencies as low as desired♣, and frequencies as high as they come; whether the upper cutoff-frequency is found to be 100 MHz or 200 MHz is immaterial in the present context. On the other hand, to deduce that all tubes would sound the same is incorrect as well. It's not that the tube itself would have a "sound", but it does influence the transmission behavior of the power stage as a whole. It does make a difference to the loudspeaker whether it is driven by a source of high or low impedance, and the character of the distortion is tube-specific, as well. The generally publicized view seems to be: tubes will sound somehow, expensive tubes will sound better, and old tubes will sound best.

Cheapest are so-called *industrial tubes* i.e. tubes manufactured for industry. Well – of course it's not only industry that gets them, because how else would they be offered in minimal quantities to musicians. "Industrial tube" probably is supposed to indicate that the musician will receive these tubes in the same condition that industry would receive them: without additional value added by the retailer. Without added value does not mean without an add-on to the price tag, though – that a business makes money from this commodity, too, is the legitimate result of mercantile aspiration. Besides industrial tubes, there are *selected/matched* power tubes. They carry mysterious numbers on their sockets and/or on their carton, and they were "paired". At least they are being boxed with a label indicating that. That such an ado will cost extra is again the result of mercantile operation. A set of 4 EL84, for example, will cost 30 Euro if you ask for industry tubes but set you back 70 Euro if you are being handed a "matched quartet". How this "matching" is done will normally not be disclosed. How well it works out: that shall be the subject of the following pages. For those of us who regard 70 Euro as an insult to their virtuosity, **NOS-ware** is available. These would be tubes that have not only miraculously be hidden away in basements and warehouses but actually were even able to reproduce, and are offered – since many years – with the supplemental encouraging remark: one of the very last originals! Their sound is portrayed as unrivaled, this assessment being supported by the intuitively fair enough reasoning that the old tube experts were scrapped together with the old manufacturing plants. In individual cases that may have been accurate (while not entirely trivial, after all), but it is – frankly – nonsense to conclude that a tube would be better just because it has spent 50 years in the basement without use. It will *possibly* deliver exactly the desired sound; just as possibly it will, however, sound worse than a low-cost industrial tube. You will only know after you've bought it.

It is difficult for the buyer to verify whether a particular tube indeed hails from ancient stock or is merely a modern el-cheapo imitation. Internet-forums about "faked tubes" are of some help here. Whether a tube does meet the given requirements will be revealed (subjectively) by listening tests and (objectively) by measuring its data. At this point we shall not yet investigate to which extent a conclusion from one to the other is legitimate – let's look at the technical data first. According to conventional wisdom, most important are plate dissipation and transconductance (plus of course the socket needs to fit). Plate dissipation = maximum load (e.g. 30 W), transconductance = gain (e.g. 5 mA/V). That, however, will not be good enough to select a power tube – in a guitar amp there are further criteria to base this choice on.

♣ Only the lifespan of the tube stands in the way of that this range not extending to exactly 0 Hz.

To **assign the power-needs** is relatively easy as long as we look at the bare essentials: low power = EL-84, 6V6-GT; medium power = 6L6-GC, 5881, KT-66, EL-34; high power = KT-88, 6550. There are of course more tubes, and some tubes were/are offered in several power categories (e.g. 6L6-GB vs. 6L6-GC), but we will not go into that here. Similarly, a discussion about the **proof voltage** will be omitted – the corresponding statements in the datasheets are too obscure and contradictory.

Power tubes are rated with about 10 – 50 W regarding the **maximum power dissipation at the plate**. This value must *not* be mistaken for the power output of the amplifier! There are 100-W-amps that draw their output power from 2xEL34 ($P_{a,\max}$ = 25W), and there are 40-W-amps using 2x6L6-GC ($P_{a,\max}$ = 30W). **Fig. 10.5.6** shows the output- and transmission-characteristics of the most important power tubes. All curves are for $U_{g1} = U_{gk} = 0$V, i.e. for full drive level. Applying positive control-grid voltages, it would in principle be possible to achieve even higher plate-currents but the usual driver stages are of too high an output-impedance for this. Besides the control-grid voltage, it is also the screen-grid voltage that determines the shape of the output characteristic. In order to be able to compare, we choose U_{g2} = 350 V, although of course not all amplifiers operate with this voltage value. The GE-datasheet even specifies as little as 285 V for the 6V6-GT – but that didn't hold back Fender to subject the 6V6-GT in the Princeton to a proud 415 V.

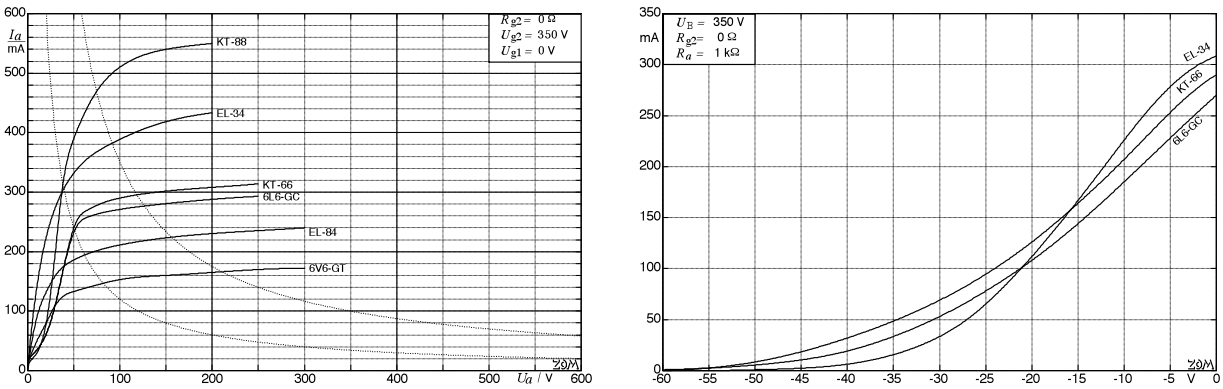


Fig. 10.5.6: Output characteristics (left) und transmission characteristics (right) of some power tubes.

We can see from Fig. 10.5.6 that – for comparable operating conditions – the maximum plate-currents differ quite substantially, after all. The transmission characteristics, as well, show pronounced individuality, and therefore a KT-66, for example, must only be switched for an EL34 after suitable modifications in the circuitry. In any case, it is important to bear in mind that such characteristics remain general, simplifying illustrations.

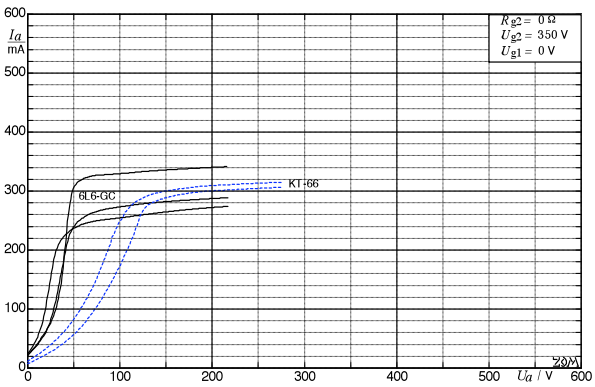


Fig. 10.5.7: Measured output characteristics

Fig. 10.5.7 proves this via measurements with 3x6L6-GC and 2xKT-66. Given just the data-sheet info, similar characteristics in all 5 tubes would be expected – reality is quite different, however. In some circles, the looks (i.e. the shape of the glass container) of a tube will be given more attention than the actual electrical function. Comparison tests that are not considering significant electrical differences as those shown above are thus not only not helpful, but just plain useless. More on that in Chapters 10.5.11 and 10.11.

The following **table** compiles some tube data. The respective year was taken from literature i.e. it does not necessarily indicate the true time when the respective tube was first issued to the market. The transconductance (mA/V) depends much on the specific operating point, and therefore the given value is for rough orientation only: detailed information is offered by the characteristic curves (Chapter 10.11).

The maximum permissible plate dissipation is also to be seen for orientation only: the specification in the datasheets of different manufacturers deviate to some extent, and moreover, back in the day the specification was done using two different standards: **Design Center System**, and **Design Maximum System** (in brackets, compare to Chapter 10.5.9).

Type	$P_{a,\max}$ / W	mA/V	Manufacturer	Year
6V6	12 (14)	4	RCA	1937
6V6-G	12 (14)	4	RCA	1941
6V6-GT	12 (14)	4	RCA	1944
6V6-GTA	12 (14)	4	RCA	1962
6L6	19 (---)	5.3	MOV ⇒ RCA	> 1933
6L6-G	19 (---)	5.3		> 1936
6L6-GA	19 (---)	5.3		> 1943
6L6-GB	19 (22)	5.3		
6L6-WGB	20 (23)	5.3	Tung-Sol	1950
6L6-GC	--- (30)	5.3		1954
5881	23 (---)	5.3	Tung-Sol	1950
7027	25 (---)	6	RCA	1958
7027-A	--- (35)	6	RCA	1959
6550	35 (---)	11	RCA	1962
6550-A	--- (42)	11	GE	1972
KT-66	--- (25)	6.3	Marconi	1956 (> 1937)
KT-66	--- (25)	7	MOV	1977
KT-77	--- (25)	11	MOV	1977
KT-88	--- (35)	11	MOV, GEC	1957
KT-90	--- (45)	11	Ei	
EL84	12 (---)	11	Philips	Ca. 1955
EL34	25 (---)	12	Philips	Ca. 1952
EL51	45	11	Philips	1953
EL151	60	13	Telefunken	1943
QB3.5/750	250	4	Philips	

Table: Power-tube data-sheet information: maximum plate dissipation and transconductance.

10.5.2 Push-pull class-A operation

The single-ended power stage introduced in Chapter 10.5.1 turned out to be relatively weak in terms of power delivery: With a 12-W-tube we could get at most 6 W output power from it. For a greater output power more powerful tubes would be available, but then there is another disadvantage of the single-ended circuit: even without any drive signal, a relatively strong DC-current runs through the output transformer, and the latter needs to operate under unfavorable conditions due to the resulting DC-**pre-magnetization**. We could insert an air gap into the iron core of the transformer and reduce the DC-field dependency of the reversible permeability – but then we would in total reduce this permeability to a value smaller than the one for the core without air gap. Moreover we need to consider that the load on the power supply is independent of the drive levels for the single-ended class-A power stage. In other words, even at rest, the power supply experiences maximum load, and therefore the supply voltage is not constant but oscillates around a mean value with a frequency of 100 Hz (given a two-way rectifier). This AC-component generates an AC-current through the output transformer, the output tube being of high impedance but still no ideal current source. The result is an undesirable interference tone at 100 Hz or 120 Hz (depending on the local power).

Using a push-pull class-A circuit (**Fig. 10.5.8**), some of the disadvantages of the single-ended class-A circuit can be avoided. The term "push-pull" is derived from the opposite-phase grid-drive of the two output tubes. The rising grid-voltage at one tube increases its plate-current while at the same time the decreasing grid-voltage at the other tube reduces the plate-current there. Ideally, the former plate-current increases by the same ΔI that the latter plate-current decreases by; the sum of the currents sourced from the power supply $I_=$ remains constant (DC current), independently of the drive levels. At rest, this DC-current splits up into the two plate-currents of equal strength that each generates a magnetic DC-field in the transformer core. Since the two DC-fields have opposite directions, they compensate each other within the core, and the latter remains field-free (without pre-magnetization). No **air gap** is necessary. A corresponding compensation also happens for the residual ripple in the supply voltage: the 100-Hz AC-current generated by it causes opposite-phase AC-fields that cancel each other out, and cannot result in **hum in the loudspeaker**.

An entirely different situation exists for the AC-currents at the plate that are created by the opposite-phase **grid-drive**: they are, in terms of the reference-arrows defined in Fig. 10.5.8, of opposite phase, but therefore correspond in-phase to the primary AC-current (defined in *one and the same* direction): $I_{aa} = I_{a1\approx} = -I_{a2\approx}$. The equality of these two AC-currents also results from the power supply (ideally) delivering pure DC: if no AC-current is leaving the winding at the tap of the primary winding, both primary AC-currents need to be equal. Assuming an ideal transformer with identical primary windings, the total primary voltage will be double the AC voltages at the plate; the two tubes thus operate *in series*.

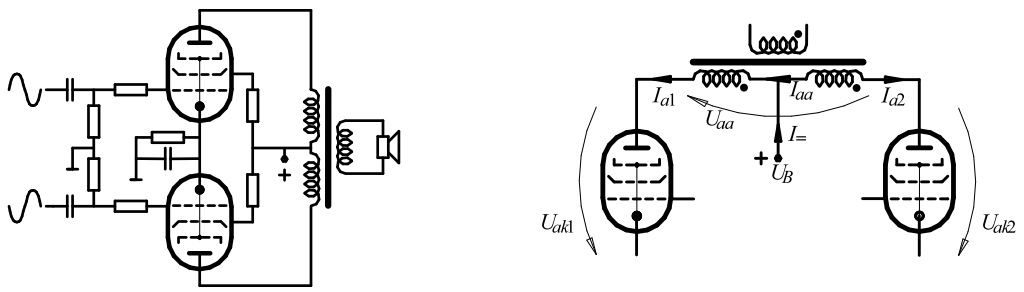


Fig. 10.5.8: Push-pull class-A power stage; on the right, the phases of currents and voltages are illustrated.

In **Fig. 10.5.9** we see the idealized voltages and currents relating to Fig. 10.5.8. Each individual tube operates in single-ended class-A mode with the operating point located in the middle of the useable load line (Chapter 10.5.1). The two tubes are driven by opposite-phase signals. With the reference-arrows defined according to Fig. 10.5.8, both the AC-voltages at the plate and the AC-currents at the plate are of opposite phase, as well. The (overall) primary voltage U_{aa} is the difference between these opposite-phase AC-voltages $U_{aa} = U_{ak2} - U_{ak1}$, and the AC-current flowing through both primary windings is $I_{aa} = I_{a1\approx} = -I_{a2\approx}$.

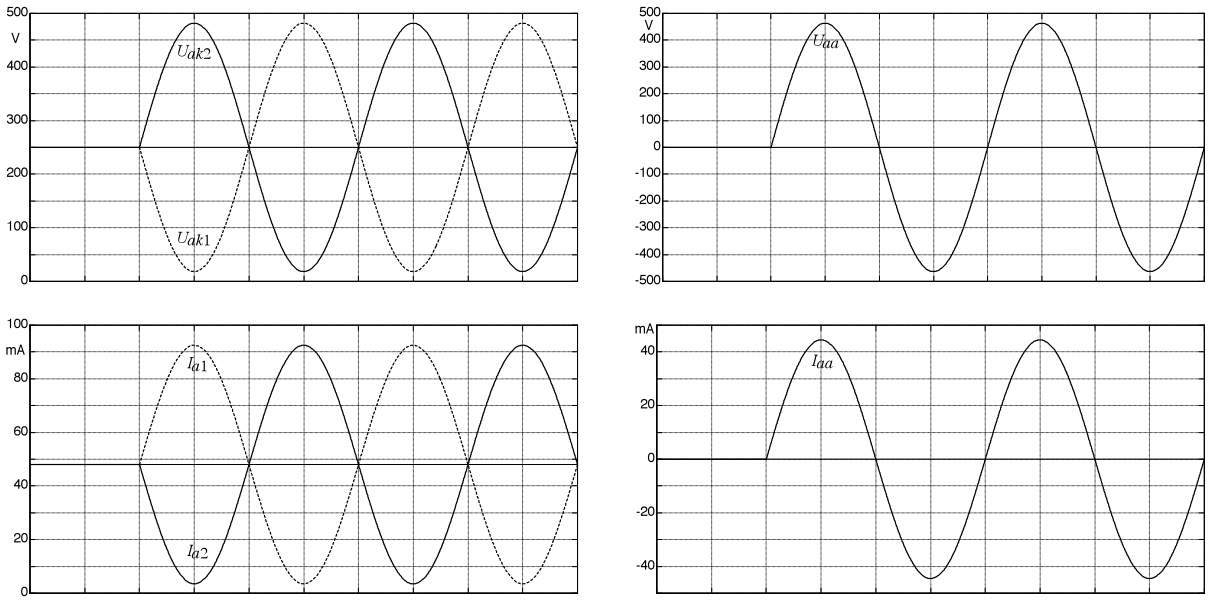


Fig. 10.5.9: Voltages and currents of the individual tubes (left); primary AC-voltage and -current (right)

If we neglect the residual plate-voltage (for $U_{g1} = 0$) and assume an ideal output transformer, the obtainable **effective power** corresponds to the maximum power dissipation in one tube; two EL84 will thus yield 12 W at the most (and in the ideal case). In practice (i.e. considering residual voltage and transformer losses), about 10 W may be expected. Ideally, the power taken from the power supply is independent of the drive level and corresponds to the maximum dissipation of both power tubes – for 2 x EL84 that would be 24 W due to the plate-currents, plus about 2.8 W screen-grid dissipation, plus about 0.8 W dissipation in the joint cathode resistor. This resistor should be chosen such that in idle the cathode-current (0.11 A in this example) generates just the required grid-bias (7.3 V).

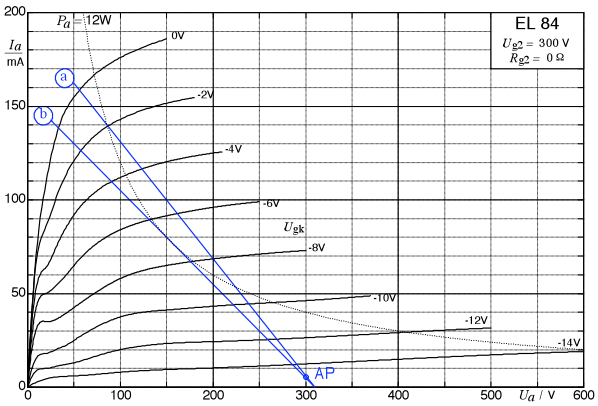
The optimum load-impedance at the plate, and with it the transformation ratio of the transformer, is to be derived from the gradient of the load line, just as it would have to be for a single-ended class-A amplifier. For both the latter and the push-pull class-A amp, every tube needs to "see" the same load impedance R_{opt}. When designing the push-pull class-A amplifier, consideration needs to be given to the fact that both (half) primary windings "see" two load impedances: the secondary winding as passive load, and the other (half) primary winding as **active load**! For this reason, the load of the individual tube in the push-pull class-A circuit may not be simply calculated from the transformation ratio (**impedance paradox**, Chapter 10.5.5)! If the transformation ratio for a single-ended class-A amplifier is e.g. $N_p : N_s = 24 : 1$, it will be $N_{p1} : N_{p2} : N_s = 17 : 17 : 1$ for the push-pull class-A amp (given otherwise equal conditions). The datasheet specifies an optimum load-impedance at the plate of $R_{opt} = 5.2$ kΩ for single-ended class-A operation of an EL84-amplifier with $U_B = 250$ V, and thus the (overall) primary impedance for the push-pull class-A amp amounts to $R_{aa} = 10.4$ kΩ.

Examples for specific amplifiers are presented at the end of the chapter.

10.5.3 Push-pull class-B operation

In a push-pull class-B amplifier, the operating point is not positioned in the middle of the load line but at its lower end. Without a drive signal, only a small idle-current flows though the power tubes. There is lesser load on the power supply and the tubes do not get as hot – however the obtainable output power is still higher than that of the push-pull class-A amplifier.

To keep the plate current small while there is no drive signal, the **grid bias voltage** needs to be on the rather negative side. Due to the small current, this cannot be achieved as a voltage drop across a cathode resistor anymore, and therefore both cathodes are set to ground potential while a separate DC voltage-source generates the required negative bias-voltage (measuring, after all, in the order of -15 … -65 V) at the grid. This DC voltage-source is designated U_{g1} in **Fig. 10.5.10**, and it is fed to the circuit via two high-impedance resistors (e.g. 220 kΩ) connected across, and two grid resistors (e.g. 1 … 5 kΩ; there are circuits without these grid resistors, as well).



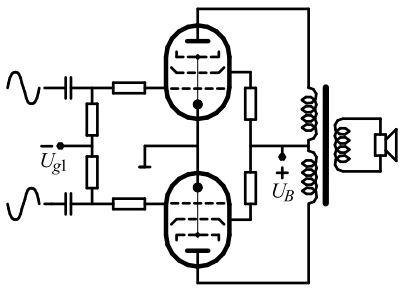


Fig. 10.5.10: Output characteristics of the EL84; circuit of the push-pull class-B power stage.

We obtain the optimum gradient of the load line (and thus the optimum **load impedance**) if the intersection of the $U_{gk} = 0$V-characteristic and the load line has the maximum distance from the operating point. **Fig. 10.5.10** shows two different load lines: the flatter line (b) relates to a 2.0-kΩ load-impedance while for (a), the load-impedance is 1.6 kΩ. It is of no issue that the hyperbola designating the power limit is intersected: the tube is only subject to strain during one half-wave, and the plate dissipation remains within the tolerable limit on average. Static drive signals, or drive signals of extremely low frequency are not to be expected with guitar amplifiers since the power stages are fed via a high-pass.

The maximum **power** yield does not differ much between the two variants: it is about 19 W for (a), and 18 W for (b). As a comparison: a corresponding push-pull class-A power stage could only deliver about 11 W. Besides the maximum power yield it is, however, also the power required from the power supply that merits consideration, especially in the case when there are small drive signals. For the push-pull class-A power stage, the load on the power supply is independent of the drive level, e.g. 24 W for 2xEL84. In contrast, the power supply needs to deliver as little as 3 W in the push-pull class-B power stage (depending on the bias setting). **Fig. 10.5.11** shows the power balance – albeit without considering the power dissipated in the screen grid that would amount to about an additional 3 W at full drive levels. In the class-B mode, the output power is larger and the power losses in the tubes are smaller: the maximum plate dissipation in class-B mode is only about half of that found in class-A mode.

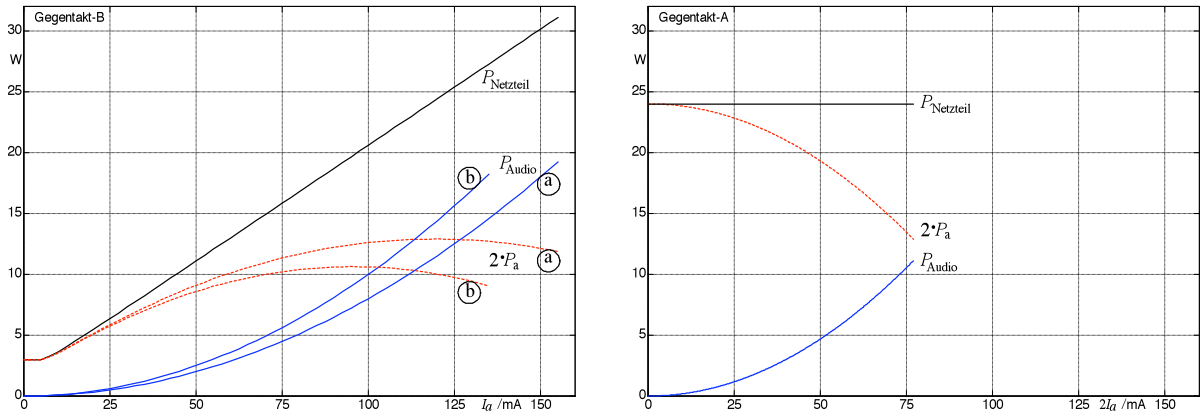


Fig. 10.5.11: Link between power-supply load (without g_2-dissipation) and output power P_{Audio}. "Gegentakt" = push-pull. The difference between the two curves corresponds to the plate dissipation $2 \cdot P_a$ of both tubes. (U_B = 300V).

The relatively high efficiency of the push-pull class-B circuit results from the fact that each tube carries a large plate current only when power is actually delivered to the load. For this, the operating point needs to be set at the lower end of the load line. However, the bias-voltage at the grid must not become too negative because this would result in **crossover distortion** (**Fig. 10.5.12**). Given a sufficiently large bias-current (left-hand section of the figure), the two tube characteristics superimpose to a reasonably smooth curve, while for too small a bias-current a saddle point appears (middle and right-hand sections of the figure). This saddle point will increase the odd-order distortion on one hand, and on the other hand leads to an undesirable (progressive) drive-dependency of the slope of the characteristic (Chapter 10.5.8). Special consideration needs to be given to the fact that the supply voltage decreases with increasing drive levels – the **screen-grid voltage** therefore decreases as well, and this further emphasizes distortion (Chapter 10.5.8).

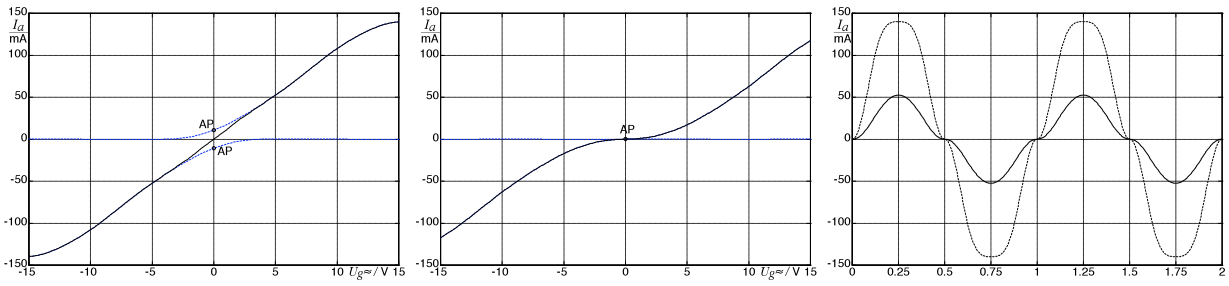


Fig. 10.5.12: Characteristics at different bias-current settings. Crossover distortion. On the right, the distortion relating to the middle picture is depicted for two different amplitudes.

Literature does **not** give an **exact definition** for the load line in push-pull class-B operation. Rather, it mentions "small plate-current", and occasionally even a plate-current for which the operating point is set "almost to zero". This did not keep Siemens and other tube manufacturers from specifying a bias current for the EL34 of no less than 35 mA. They do have a point because the theoretical case that the plate-current approaches "almost zero" has next to no bearing on low-frequency applications. 35 mA: that is indisputably "somewhat more than almost zero", and whether this mode of operation may in fact be still called "push-pull class-B" is subject of controversial discussions. Alternatively, the term "push-pull class-AB operation" is used, or the term "push-pull class-D operation" – it is important to know that these designations are ambiguous! (Details are found in Chapter 10.5.4).

In class-B operation, the two power tubes conduct simultaneously only for small drive levels; at higher drive settings each power tube conducts predominantly only during one half-wave. This needs to be taken into account when choosing the **cross-section of the wire** in the transformer. If we assume a sinusoidal drive signal, and a **peak plate-current** of 141 mA (Fig. 10.5.10), the RMS-value is not 100 mA but only 50 mA.

The **plate-voltage** without drive signal is, for class-B operation, just slightly less that the supply voltage U_B (e.g. 300 V). During the half-wave at which the tube is conducting, the plate-voltage drops as far as the residual voltage (e.g. 30 V). During the other half-wave (blocked mode), the plate-voltage does not remain at the level of the supply voltage but increases to almost double of that value (e.g. 570 V!) This is because the primary winding of the output transformer sees practically no load when the respective tube is in blocking mode, while the magnetic flux generated by the other (active) tube induces a high voltage in this winding without load. In power stages that operate with higher supply voltages, voltages that are even much more dangerous can result: e.g. 850 V in Fender amplifiers, or 1100 V in Marshall 200-W-amps. These high voltages are not contradictory to the information given by datasheets where the maximum plate-voltage is specified e.g. at 800 V; the values expressed there are meant as idle voltage (without drive signal). For example, the datasheet of the EL34 determines the maximum plate-voltage at 800 V but allows for maximum peak voltages of 2000 V in the blocked mode. Such high voltages can in fact occur easily if the amplifier is not connected to its nominal load but operated with a higher impedance, or no load at all at the output. In this case, spark-over or arcing between the connector-pins 3 and 2 (plate and heater filament) can easily happen – which is likely to damage the tube socket and/or the tube holder irreversibly. Even more dangerous is an insulation-destroying spark-over within the output transformer because an adequate replacement for this component may not be at hand.

A few comments regarding seemingly “useless” circuits-components: that they are included often needs to be credited to practical insights. The grid-resistors (2 – 5 kΩ) connected in series with the (apparently high-impedance) control grid will reduce the tendency of a power stage to self-oscillate. High-frequency self-oscillations may occur – but they do not have to. The power stage may well operate perfectly without these resistors, as well; however, it is advisable not to simply omit them. With each tube- or loudspeaker-change, different stability-criteria creep in, and the small additional investment for these resistors can very quickly pay off. The same holds for small capacitors (10 – 100 pF): if they are not directly connected to the tone-filter stages, they presumably are supposed to suppress RF-oscillations. It is indeed possibly that they were (had to be?) chosen with a value that audibly cuts into the brilliance of the guitar sound. If that is the case, we find a wide field of possibilities to improve the sound – but we are also confronted with a good chance that we operate a powerful radio-transmitter as we change or remove such capacitors. Since power-stage oscillations can easily occur in the FM-range (100 MHz), it is recommended to check the stability with a broadband oscilloscope. Evidently, we must not discard such oscillations as “inaudible” and therefore irrelevant: on one hand, operating such a transmitter may be unlawful, and on the other hand the power tubes may be overloaded massively. Moreover, there may well be secondary symptoms that are audible.

10.5.4 Push-pull class-AB operation, push-pull class-D

A push-pull class-A power stage operates in push-pull class-A mode for small drive levels, and for high drive levels in push-pull class-B mode – that far, literature agrees. In detail, however, differences appear and we find three definitions that we will designate *old*, *alternative*, and *new*. According to the **old definition**, the class-AB operation is a class-B operation with a somewhat enlarged bias-current; there is a distinction between AB_1 (without grid-current) and AB_2 (with grid-current). A specific guideline where to set the operating point does not exist; it may be located (in the output characteristic) "somewhere between" the A-operating point and the B-operating point. This has often led to defining the location of the AB-operating point exactly in the middle between the two (A- and B-) operating points. Example: if the bias-current is 50 mA for class-A operation and 10 mA for class-B operation, then it must be 30 mA for class-AB (according to the datasheet).

The literature from "back in the day" does not specify whether the bias-voltage at the grid of the class-AB circuit is generated "automatically" via a resistor at the cathode, or via a separate voltage source. The **alternative definition** seeks to be more precise. In the class-AB amplifier, the operating point can shift dependent on the input signal: with increasing drive level, the cathode-current will become more and more asymmetric (due to the non-linearity of the characteristic). Consequently, the voltage-drop across the cathode resistor (bridged by a capacitor) increases and shifts the average grid/cathode-voltage more into the negative. That way, class-A operation changes into class-B operation as drive levels increase (**Fig. 10.5.13**). The alternative definition moreover designates all those power stages with the term **push-pull class-D amplifier** that generate their bias voltage at the grid (exclusively) via a separate voltage source, and that feature an increased bias current relative to the class-B operation [e.g. H. Schröder, W. Knobloch]. This definition does not generally consider the polarizations of the coupling capacitor also leading to a drive-level dependent shift of the operating point.

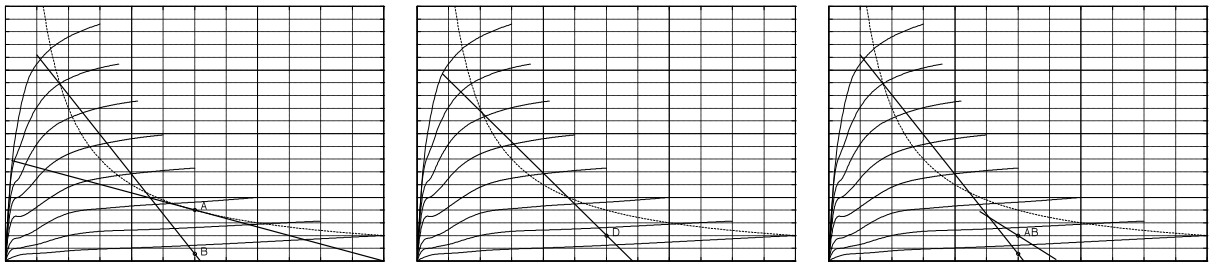


Fig. 10.5.13: Output characteristics and operating point: push-pull class-A and –B amplifiers (left), push-pull class-D amplifier according to the alternative definition (center), push-pull class-AB amplifier according to the alternative definition (right).

Under the moniker class-D operation, the **new definition** considers something entirely different: it designates a switching amplifier (using pulse-width modulation PWM) with the term D. According to the new definition, class-AB operation is a class-B operation with increased bias-current and a fixed operating point. Presumably, this "new" terminology came in when bipolar transistors started to supersede power tubes. Setting the operating point for transistor circuits is done according to different criteria compared to tubes; there is no drive-dependent operating point anymore, and the meaning of the terms changed.

In contrast to HiFi power amplifiers, the minimization of distortion does not have priority in typical guitar amps. For this reason, we see a domination of old-school class-AB power amps with the bias-current set according to special criteria (Chapter 10.5.8). The AC-30 also belongs to this group, and *not* to the group of push-pull class-A circuits (Chapter 10.5.12).

10.5.5 The impedance-paradox

The output transformer matches the low-impedance loudspeaker (e.g. 8 Ω) to the higher-impedance tube circuitry. For push-pull circuits, this transformer has *two* serially connected primary windings. Could you say what the input impedance of these two windings is? Let's take as an example turns-ratios of 10:10:1, and 8Ω as secondary load: for the ideal transformer the input impedance of the whole primary winding is $R_{aa} = 20^2 \cdot 8\ \Omega = \boldsymbol{R_{aa} = 3200\ \Omega}$. Is now the input impedance of one of the two windings half of this value i.e. **1600 Ω**? For the push-pull class-A operation, we assume as much because here the same AC-current flows through both primary windings. However: calculating the impedance transformation for half the primary winding, we get: $R_a = 10^2 \cdot 8\ \Omega = \mathbf{800\ \Omega}$. What is the correct value?

The push-pull transformer is a **three-port network** i.e. a system with three pairs of connections. The two primary ports are connected in series so that overall only 5 connecting points show. To calculate the input impedance of one port, the two other ports need to be considered as loads. If we connect only *one* 8-Ω-load resistor to the secondary winding, and leave the one primary winding open, we will measure 800 Ω at the remaining other primary winding. This is because the transformer does now operate merely as a two-port network (= quadripole). However, if we connect *both* primary windings – as it is done for the push-pull class-A operation, then each primary winding "sees" *two* load impedances: the secondary load, and the other primary winding.

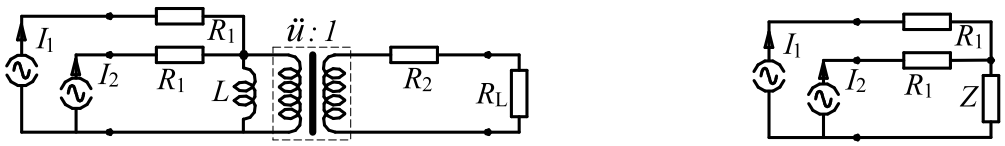


Fig. 10.5.14: The rigidly-coupled output transformer as three-port. Right: simplified circuitry; *ü* = *TR*.

Fig. 10.5.14 shows a simplified equivalent circuit of a transformer. R_1 and R_2 are the resistances of the windings, R_L is the secondary load resistance, L stands of the main inductance. In the middle frequency range the effect of the main inductance may be neglected, and the secondary resistances can be transformed via $(TR)^2$ into the equivalent impedance Z. The equivalent circuit given in the right-hand section of the figure can now easily be calculated:

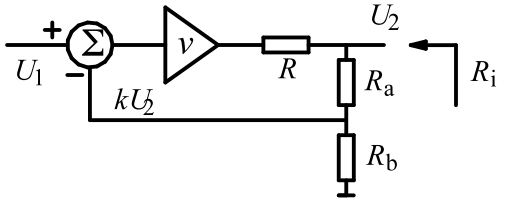
$$Z_1 = R_1 + (1 + I_2 / I_1) \cdot Z \qquad\qquad Z_2 = R_1 + (1 + I_1 / I_2) \cdot Z$$

It is clear that for $I_2 = -I_1$ the input impedance becomes independent of Z (or rather of R_L), because here the voltage across Z approaches zero. In a push-pull class-A power stage, however, the two currents are in opposite phase (and ideally also equal in their magnitude) so that the input impedance is increased. Neglecting R_1, the input impedance doubles as we bring the second primary winding into the circuitry.

Let us include some numbers into the above **example**:
In the **push-pull class-A power stage**, the input impedance of (half of) the primary winding is half of the input impedance R_{aa} of the total primary winding, i.e. 1600 Ω. In contrast, only one winding is active at any given time in the push-pull class-B power stage: when one of the tubes is conducting, the other blocks. Therefore, in this case the input impedance of (half of) the primary winding is only a quarter of R_{aa}, i.e. 800 Ω. The impedance R_{aa} does not appear physically in the push-pull class-B power stage; it remains a pure calculation value.

10.5.6 Negative feedback

We talk about **feedback** if a part of the output signal of an amplifier is channeled back to the input and superimposed there onto the input signal. Same-phase feedback is termed **positive feedback** while the designation of opposite-phase feedback is **negative feedback** (sometimes abbreviated with NFB in the following). Since there are two output signals (current and voltage), and correspondingly two input signals, four different ways of negative feedback may be defined. In a typical guitar-amp power-stage we predominantly find negative feedback of the voltage-voltage kind: a percentage of the output *voltage* is fed back and superimposed on the input *voltage*. This superposition results in a **control circuit**: as the output voltage decreases (e.g. due to loading), less voltage is fed back – resulting in more gain so that the voltage loss is partially compensated. This special negative feedback (termed g_{21}-negative-feedback in circuit-design) stabilizes the voltage-gain factor and reduces the linear internal impedance♣, and also broadens the small-signal-bandwidth, and reduces harmonic distortion.



$$R_i = (R_a + R_b) \,//\, R_{gk}, \quad R_{gk} = R/(1+kv)$$

$$v_U = U_2 / U_1 = \frac{v}{1 + kv + R/R_L}$$

Fig. 10.5.15: Basic schematic of a feedback loop (left). For positive gain v, negative feedback results. The most important values are the internal impedance R_i and the voltage-gain-factor v_U.

In **Fig. 10.5.15**, k determines the degree of the negative feedback i.e. its effectiveness. For k = 0, the negative feedback is without effect; with rising k, the effect of the negative feedback increases. R represents the internal impedance of the power stage without feedback; in tube amplifiers, this is considerably larger than the load impedance. The **Fender Super Reverb**, for example, reveals $R_i = R = 180\ \Omega$, and $v = 160$; with a load of 8 Ω[1] (R_L) the voltage gain will be $v_U = 6.8$. The factory-set negative feedback is $k = 0.056$; with it R drops to $R_i = 18\ \Omega$, and the gain to $v_U = 4.9$ (measurement: **Fig. 10.5.16**). The low-frequency range reveals an interesting twist: due to phase-shifts, a positive feedback comes into play here! Enlarging the input capacitor of the differential amplifier (from 1 nF to e.g. 100 nF) will, however, keep the output impedance in the low-frequency region (with NFB) almost constant (Chapter 10.4.3).

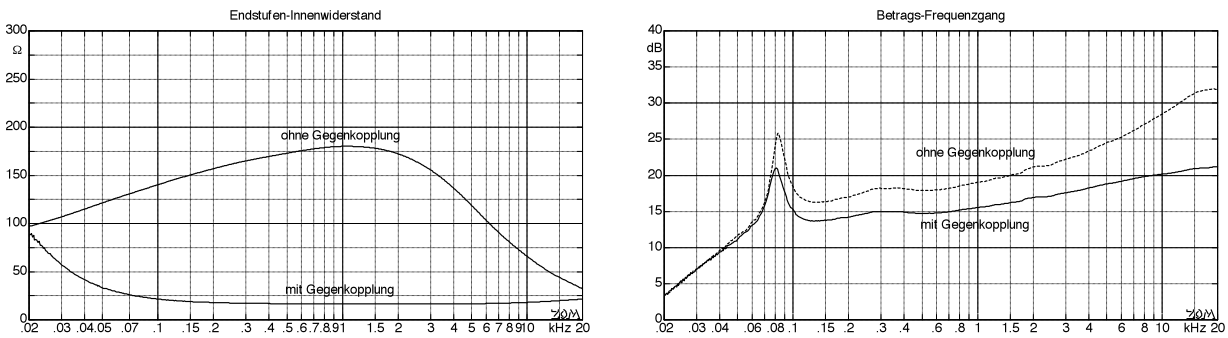


Fig. 10.5.16: Left = magnitude of the output impedance. Right = frequency-response, from the input of the differential amplifier (ahead of the 1-nF-capacitor) to the loudspeaker (4xP10-R).
"Endstufen-Innenwiderstand" = power stage impedance; "Betrags-Frequenzgang" = magnitude frequ. fesponse;
"ohne Gegenkopplung" = without NFB; "mit Gegenkopplung" = with NFB.

♣ For non-linear operation (overdrive), any negative feedback will loose its effect since the control value (here: output voltage) can practically not change anymore.

[1] Note that the Super-Reverb-specimen investigated here had an output transformer with not just the customary 2-Ω-output, but also an additional output for 8-Ω speaker-matching.

Not all power stages include negative feedback: the VOX AC-15 and AC-30 (and some others, too) completely dispense with any feedback. In Fender amps, the situation is mixed: very early power stages have no feedback, the Bassman 5B6 receives NFB around 1952, the Deluxe 5E3 only as late as 1955 [D. Funk]. The Bassman acquires special significance due to its Presence-control included in the negative-feedback circuit: it enables the frequency dependency of the feedback to be set by the user (Chapter 10.3.3). Legend has it that Jim Marshall and Ken Bran were particularly inspired by the late 1950's Fender Bassman when they developed their amps, and therefore we find a power stage with integrated presence potentiometer in Marshall amps, as well.

As early as 1943, Frederick Terman describes, in his remarkable "Radio Engineers' Handbook", the effects of power-stage feedback on gain, internal impedance, and harmonic distortion. The reduction in gain that accompanies negative feedback certainly was not an express objective of the circuit designers, but they put up with it in order to reduce the non-linear distortion of tubes and output transformers. In the 1950's, there was no Heavy-Metal music scene, and playing was mostly "civilized" i.e. undistorted. Presumably, the pioneering developers observed the output signal of their amplifiers with an oscilloscope, and tried to reproduce sine curves as perfectly as possible: "*by the book*", as D. Funk writes. The more negative feedback is introduced, the less an amplifier distorts – that's what the book said. It was also known that strong phase shifts may turn NFB into positive feedback – although not every designer would or could do much of the required calculating. In any case, the designer would soon discover that, with too strong a negative feedback, the amp would start to self-oscillate, and so the NFB was adjusted empirically to a degree that would avoid instability within the framework of the given manufacturing tolerances.

In **Fig. 10.5.17**, we see the frequency responses of a bandwidth-limited system: in the left-hand section for slightly different filter flanks, and with and without negative feedback. The more narrow-band system (dashed lines) does not only receive the expected gain-reduction but also considerable resonance peaks at the frequency limits. "Negative feedback" means superposition of an opposite-phase signal. However, the phase-shifts that live in every circuit with bandwidth-limiting will have the effect that around the band-limits, the opposite-phase correction-signal can turn into an (almost) same-phase positive-feedback-signal that increases gain. Increasing the gain in the forward branch (right-hand section of the figure) will disproportionally increase the overall gain (blue) in the fringe ranges. In fact, this may occur as an effect of just a tube change. The new power-stage tubes will generate a stronger bass response due to their slightly higher transconductance, and right away the test report in the music-mag will read: "the KT-X delivers more bass than the 6L-Y". This characteristic, however, needs to be always seen in connection with the specific individual circuit. Power tubes will transmit from 0 Hz to about 100 MHz – but only the teamwork also including transformer, speaker-load and NFB-network will result in the individual frequency response!

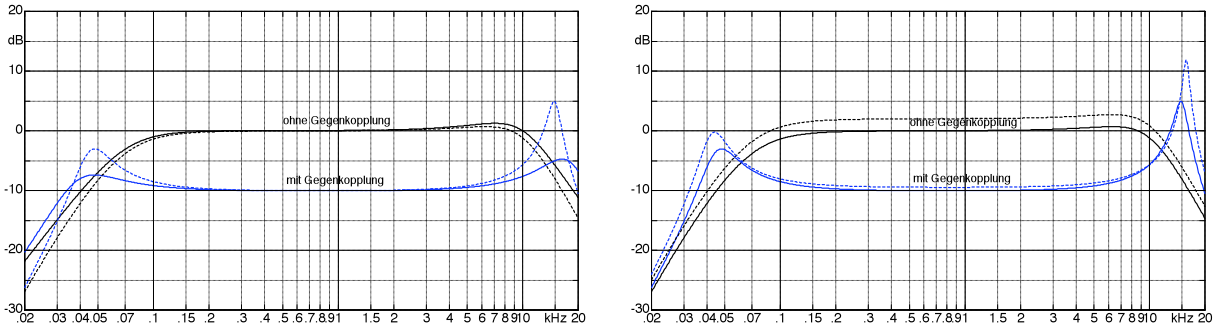


Fig. 10.5.17: Effects of negative feedback when a bandwidth-limitation is present. "Gegenkopplung" = NFB; "mit" = with; "ohne" = without.

10.5.7 Internal impedance of the power stage

Tube circuits are of high impedance while loudspeakers have a low impedance. The output transformer – with the term "matching" appearing in its description – serves as mediator between these different impedance levels: indeed, the output transformer *matches* the different impedance levels *to each other*. Usually, the term "matching" indicates that source- and load-impedance are of equal magnitude. For a tube amplifier, this would mean that its internal impedance is decreased to the level of the loudspeaker by the output transformer, i.e. is brought down to e.g. 8 Ω. This strict definition of the term "**matching**" should, however, not be used in the context of tube amplifiers; the impedance levels are brought closer to each other, but they are not actually (power-) matched. In a tube amplifier specified for a nominal loudspeaker-impedance of 8 Ω, the internal impedance of the amplifier (the source impedance) will normally not be 8 Ω but much more, e.g. 100 Ω. Tube amplifiers operate "almost as current sources", and not with power-matching.

For HiFi-loudspeakers, such a current imprint is usually unwanted because an emphasizing of the loudspeaker resonance will result. Customary is the operation from a source with very low impedance that reduces the undesired high Q-factor [3]. This ideal can be achieved almost perfectly with transistor power-stages, while in tube power-stages, the impedance can be reduced via negative feedback – but not to the same degree as in transistor amplifiers (phase-shifts, tendency to self-oscillate). The question whether **negative feedback** is at all desirable in **guitar amplifiers** has received quite different answers in the past: no negative feedback in almost all VOX amplifiers and very early Fender amps; inclusion of negative feedback in almost all Fender amps from the early 1950's. The amplifier with NFB reacts "more civilized" with lower non-linear distortion compared to its feedback-free counterpart – at least as long as it is not overdriven. Whether the lower distortion is felt to be an advantage is a matter of taste and shall not be the subject of an evaluation here. However, since the negative feedback does not only influence harmonic distortion and dampening of the loudspeaker, but also has an effect on the source impedance of the guitar amplifier, another question becomes obvious: can the **output power** be increased via negative feedback? The NFB as we see it in power stages does decrease the ("too high") internal impedance of the amp – it should be possible to interpret this aspect as an improvement of the power-matching situation.

Things are not that simple, though – the tube is a **non-linear component** that is only inadequately described by the theory of linear two-ports. Experience shows that it is conducive to distinguish between (approximately) linear and (strongly) non-linear operation. For small drive-levels, the power stages works approximately in a linear fashion♣. In this case the internal impedance of the tubes can be estimated from the slope of the output characteristic. Depending on the type of tube and on the operating point, we can expect an internal tube-impedance of 10 – 100 kΩ. If we would now chose – in order to achieve power matching – the load resistor at the plate exactly as big as the internal impedance of the tube (e.g. 100 kΩ), then the AC plate-voltage would have to be 3.1 kV_{ss} in order to reach P = 12 W … no normal tube could withstand that. Equal impedance definitely is not the desired goal; rather, the output power is to be maximized while considering the given limit values. In the chapters on the specific push-pull power stages, we will give guidelines for calculating the optimal load resistor at the plate – typically, values around 1 – 2 kΩ are the result.

♣ Given that a sufficient bias-current has been set in the case of push-pull operation.

The optimal load-resistance for the plate is considerably smaller than the internal impedance of the tube, and therefore the internal impedance of the transformer is considerably larger than the nominal impedance of the loudspeaker. The loudspeaker voltage consequently depends strongly on the loudspeaker impedance. This is not a disadvantage, though, because – in contrast to HiFi-loudspeakers – emphasizing speaker- and enclosure-resonances is not generally undesired in guitar amplifiers. In fact, it is seen as a special sound characteristic and even asked for in many cases. Still, it needs to be considered that the stiff-current-source feature dies a rather sudden death as clipping occurs. The power stage is of high output impedance only while it remains in linear mode; for overdrive, the plate-voltages of the push-pull amplifiers (and therefore the output voltage of the output transformer, as well) hit a relatively rigid border: the residual voltage of the tube (e.g. 50 V). **Fig. 10.5.18** shows the output characteristic of a 6L6-GC in combination with a few internal impedances. Just like any other tube, the 6L6 does not have <u>one single</u> internal impedance. Rather, the latter is strongly dependent on the drive-signal level, and it changes by up to two orders of magnitude as the operating point shifts. The tube will be of high impedance (about 70 kΩ) in the range of usual bias currents (e.g. 350 V, 30 mA), but become of lower impedance at the overdrive-limit (e.g. 50 V, 200 mA). Of course, this is not really that surprising because the tube is a strongly non-linear component. We need to always remain aware of this, especially since the theory of LTI-systems with its relatively simple calculation methods is all too alluring – just like it is also deceptive. Connecting a 14:1-transformer to the **6L6-GC** shown in the figure, the transformation will be from 72 kΩ to 367 Ω, which is a high impedance in comparison to an 8-Ω-speaker. The transformer will, however, transform the 2 kΩ to 10 Ω.

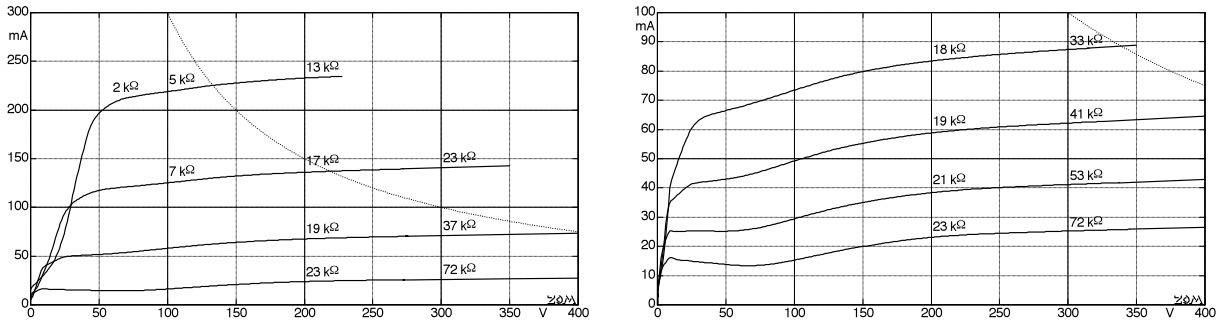


Fig. 10.5.18: Output characteristics of the 6L6-GC, including a few internal impedances. U_{g2} = 300V, R_{g2} = 0.

The internal impedance specified in the **datasheet** of a tube is an orientation-value that may be used for small-signal considerations as a rough approximation. More extensive calculations using it are not advised; first, because the power stage rarely operates under small-signal conditions, and second, because the internal impedance is specific to a given operating point, and on top of that it also depends on the voltage at the screed-grid. For the 6L6-GC, the datasheet specifies an internal impedance of 33 kΩ (class-A). This is a good match to the measurement data given above but remains usable only for very few guitar amplifiers because they normally operate in class-AB-mode. For the latter, datasheets usually do not give any internal impedance – rather, the optimum load impedance is given. This optimum load impedance – and not the internal impedance – may serve to calculate the transformation ratio of the output transformer. To calculate the source impedance R_Q (as it is "seen" by the speaker) for class-AB operation, several peculiarities need to be considered. For small drive levels, the two power tubes cooperate and R_Q is halved: with a 10:10:1 transformer, we obtain R_i = 60 kΩ → R_Q = 300 Ω. For high drive levels, only one tube is active at a time (for each half wave). Moreover, we need to consider that the transformer is not at all ideal, either: R_Q is reduced by the (non-linear!) main inductance and the capacitance of the winding.

Fig. 10.5.19 shows measurements of a power stage (JTM-45, KT-66, GZ-34). The uppermost curve results from switching-off the supply voltage (Standby): the tubes (with the heater in operation) do not insulate perfectly but are of rather high impedance – transformer-capacitance and -inductance determine the impedance. Switching on the supply voltage but keeping the negative feedback deactivated (no NFB) reduces the internal impedance R_Q because the tubes are now operated in the operating point. As we switch on the negative feedback, R_Q experiences another, very pronounced drop. Since the frequency response of the feedback loop can be modified by the Presence potentiometer, various characteristics may be realized. The strong resonance peak at 7 kHz is due to phase-shifts caused by the presence-filter (low-pass within the negative-feedback loop, see also Chapter 10.3.3). The increase in the no-NFB-condition does not happen proportionally to the frequency: this is due to the non-linear main inductance that depends on the drive levels, and on the operating point within the hysteresis (Chapter 10.6).

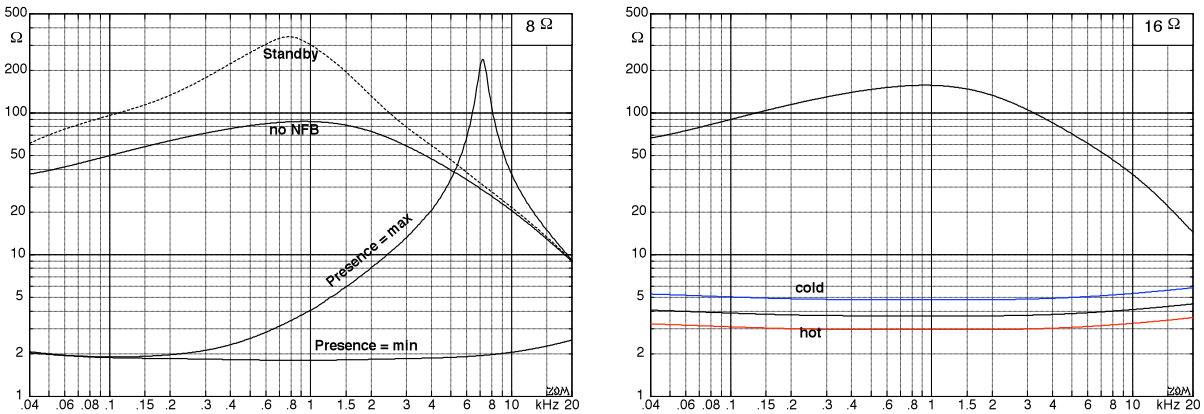


Fig. 10.5.19: Internal impedance of the JTM-45-power-stage. Left for the 8-Ω-output, right for 16 Ω. The right-hand picture also shows the effect of the bias on the internal impedance.

The impedance also depends on the bias-current of the power tubes, and on the power-tube type. If the power tubes are not equal, pre-magnetization effects of the transformer core weigh in, as well. Because of these dependencies, it is advisable to take from Fig. 10.5.19 not more than the fact that output impedances around 100 Ω occur without negative feedback. Any exact data or frequency responses would be too much connected to the individual amplifier. On the other hand, taking into account that normally the feedback-loop in the KTM-45 is closed, the differences in regular operation may not be that big, after all. With closed NFB-loop, we see an astonishingly small internal impedance (= magnitude of the output impedance) of the power stage of merely 2 Ω (for the 8-Ω-output). This power amp does have efficient NFB! Well, that's the case at least if we don't turn up the Presence-control too much … Who would have though that Marshall (not actually known for any HiFi-designs) would decrease the output impedance via negative feedback (that will decrease distortion) to values that are significantly below the load impedance!

At Marshall, this take on things would not always remain, as the 18-W-amp developed later proves. Its two EL84's operate in a power stage that entirely does without any negative feedback. VOX immediately comes to mind, but allegedly the "Watkins Dominator" was the inspiration for Ken Bran [Doyle]. Still, there are no big differences to the AC-30 with regard to the internal impedance, as Fig. 10.5.20 shows. Also generally valid: such measurement logs are snapshots – every tube-swap will change the amplifier-parameters and as such also the internal impedance. Chapter 10.5.11 will elaborate on how far selecting the power tubes helps to avoid inter-individual differences. Moreover, the effects of the bias-setting are discussed in Chapter 10.5.8.

Of the output-impedances documented in **Fig. 10.5.20**, three are taken from power stages that do not include negative feedback: VOX AC-30, Marshall 18W and Tweed Deluxe; the Super-Reverb does have NFB but was additionally measured with open negative-feedback loop.

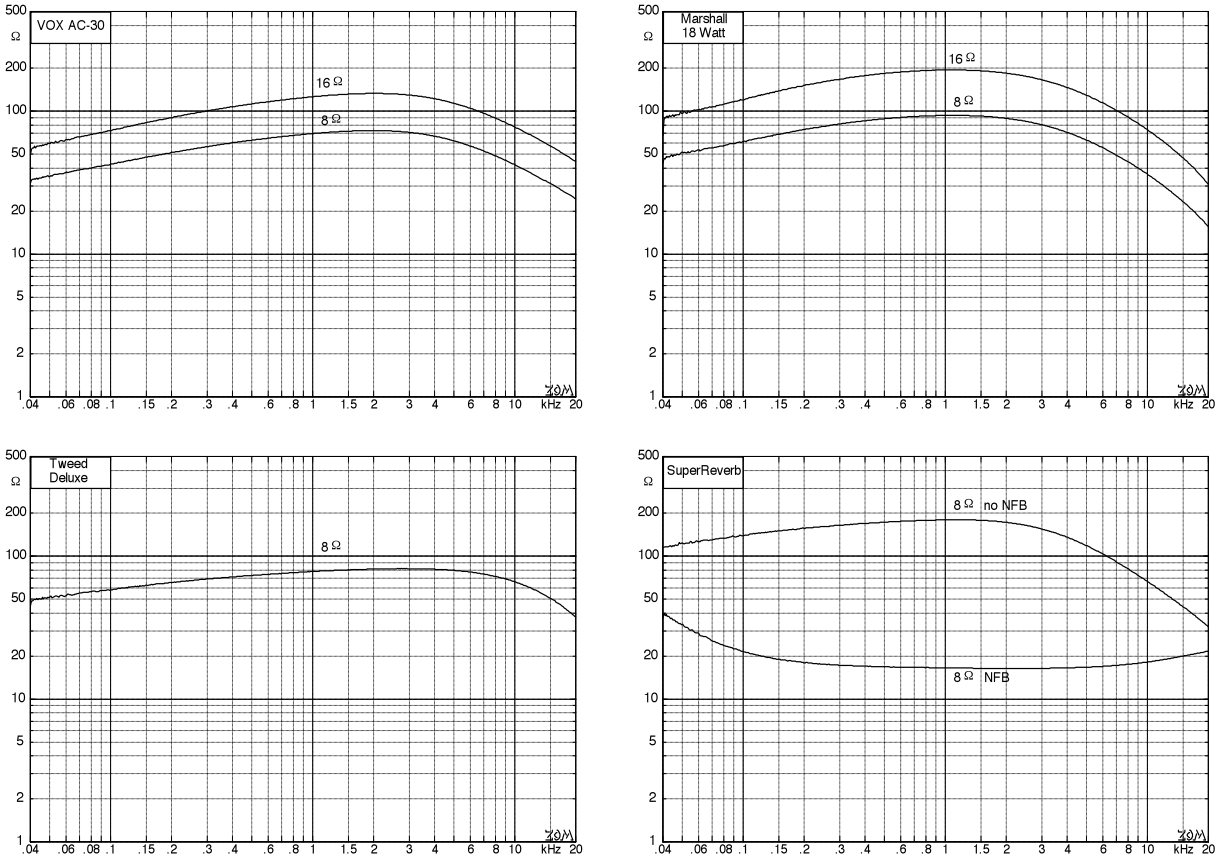


Fig. 10.5.20: Internal impedances of several guitar amplifiers (Super-Reverb with and w/out negative feedback).

In view of these significant internal impedances, we could ask where the **energy** necessary for their operation is actually sourced. If an 8-Ω-resistor absorbs 30 W and if it has serially connected an impedance of 75Ω (the internal impedance of the amplifier), will the latter then absorb 281 W? Is that why the VOX gets so hot? No – with this thinking, we would in fact abuse a model. *With respect to a specific given problem, an equivalent circuit shows the same behavior as the real structure* [20]. In our case, however, the given problem (the real source and its replacement by an ideal source including an internal impedance) is <u>not</u> the energy balance. Rather, the difference between source voltage and terminal voltage is to be illustrated. We see this right away as we replace the voltage source (with a serially-connected internal impedance) by a current source (with a parallel-connected internal impedance): with open terminals, the internal losses for the current source are at a maximum, and zero for the voltage source. The model of a source with internal impedance is well equipped to explain the dependency of the output voltage on the load impedance (loudspeaker impedance): for a low-impedance source, the terminal voltage is practically independent of the load, while for a high-impedance source, the terminal voltage is practically proportional to the load impedance. Model and reality are a good match for visualizing the given problem. However, the model is not suitable to determine dissipation power in tubes: the actual (real) voltages and current in the tubes need to be considered for this.

10.5.8 The bias-current in the power stage (bias-setting)

The bias-current in the power stage is the current that runs through a power tube when it is "in idle"-state. "In idle" means that all supply voltages are present (in contrast to the "standby"-state) but no input signal is fed to the circuit (volume = 0). In some circuits, the bias-current sets itself automatically (cathode-resistor, Chapters 10.5.2 and 10.5.2), in others it can be adjusted within certain limits by a potentiometer (bias pot). For such an adjustment, instructions are required: what is the optimum setting, and how (and what) do we measure?

What is measured? The bias, of course! *"If the mains voltage drops from 230 V to 220 V (translator's remark: or from 110V to 105 V, if your are in the corresponding part of the world), the bias changes by a few milliamps, and the sound does change"*. Of course: the bias! Adjusting that just right, the sound will be right! The bias; that's apparently the idle-current in the power stage. But which one – the cathode- or the plate-current? This is where it already gets tricky (and slippery) for most experts. Recommendation 1: *you measure the bias by disconnecting the plate from the output transformer and putting an ampere-meter in between.* This expert clearly targets your plate-current. And if, while you're measuring away, suddenly Jimi H. appears and invites for a jam: then the insulation was inadequate. Because **plate-voltages can be – SERIOUSLY! – absolutely DEADLY!** Measurements of this kind are not something the layperson can do; real expert knowledge is required. Recommendation 2: *You measure the bias by connecting an ampere-meter in parallel to the primary winding of the output transformer.* Our second expert also targets the plate-current and sees a measurement error (due to the copper resistance) of 5 – 10% as unproblematic. These primary windings are not of that high an impedance, possibly as low as 30 Ω. For an orientation measurement, this is good enough, though. Recommendation 3: *You insert (solder) a 1-Ω-resistor into the cathode-connection of every power tube and you measure the voltage drop across it.* Ooops – now we've jumped to the cathode-current, i.e. the sum of the plate- and the screen-grid-current. For a plate-current of 35 mA, the screen-grid-current may well be 5 mA with the result that the cathode-current is 40 mA. If we think of a 5%-change in the mains voltage as substantial, we should not include a 14%-error in our current measurements. Most serious datasheets specify the **plate-current** in the operating point; measurements at the cathode resistor would give us the **cathode-current**. That in fact is no problem if the screen grid (g_2) is operated with a grid-resistor in series (e.g. 470 Ω): in this case the screen-grid-current can easily be calculated from the voltage drop across this resistor. Still: CAUTION! This measurement, too, can have a deadly conclusion … the same danger that always exists when doing measurement on the opened-up amplifier. Do observe all regulations!

Instead of recommendations relating to the plate- or cathode-current in idle state, we also find hints towards an optimal setting of the bias-voltage at the grid: *adjust to -42 V at the grid (g1) of the power tube.* Indeed, this also is a workable approach: measure – using a volt-meter with high input impedance – the voltage between grid and cathode with no drive signal present: the more negative this value, the smaller the plate current, i.e. at -50 V there is less current through the tube than at -40 V. The actual current value is, however, not revealed this way.

So, what **is** the correct voltage or the correct current? Answers fill many thousand pages on the Internet; it's a science in itself. Correction: it's a playground for self-proclaimed experts, not a science as such. Searching for Ohm's law, you will consistently find $U = RI$. Looking for rules to set the bias, results are contradictory. One advice might be to use an oscilloscope and "*turn up the bias until the kinks in the curves disappear*". Plausible, that one: the spelling is correct – must be a studied person. But the next entry calls exactly this method: "*couldn't be further from the truth*". Is it even more plausible because the guy has 1532 postings?

Well then, let us add version 1001 to the 1000 existing ones. First, however, and as always, we need to suffer through some basics. In **push-pull** power stages (and only those are discussed here, anyway), the audio signal is first dissected into two parts that are amplified separately and then re-joined. The separation- and re-joining processes are error-prone, and it is here that the bias-current adjustment helps out. Changing the bias-current may improve the sound – or make it worse if you don't do it right. If the bias-current is set too low (cold biasing), distortion of the not-so-nice kind appears. At the same time the power-stage has an expander-effect: a lightly plucked string will be reproduced too softly, and with a stronger attack the amp suddenly roars. For the bias-current set high (hot biasing), the sound is good (if there are no other issues). So, should we set the bias-current as high as possible? No, don't – that will reduce the power-tube lifespan (which anyway is relatively short) even further, and could possibly destroy the power supply if it is under-sized. This would be the main effects.

In the details, 2nd-order-effects show up, as well. For small bias-currents, the filter-capacitors in the power supply get charged to a higher voltage, which might give the preamplifier- and intermediate-amplifier-stages a different operational behavior. We should not expect big effects from this but it should be mentioned for completeness sake. A small effect could also manifest itself in terms of the impulse-power i.e. the power measured at the onset of a tone. If the filter caps are charged to a higher voltage, the impulse power, too, will be a little higher. It is, however, not purposeful to reduce the bias-current just because of such effects – the distortion connected to the readjustment is normally not acceptable. If we do not start with the details but stick with the main effects, we have a simple rule: **low bias-current = distortion**, **high bias-current = premature death of the tubes**.

But then, we also find: high bias = distortion, low bias = tube-death. How can that be? Simple, actually: the experts, in particularly the self-proclaimed ones, writing (rather: allowed to write?) their columns in the guitar-magazines do have very different educations♣. The term bias is not always meant to refer to the actual bias-current but may be used as for the bias-voltage fed to the grid of the power tubes in idle. This is where a mix-up may well happen, and even a double mix-up at that, because for the negative bias-voltage, it is easy to confuse the magnitude of the given number and the actual value (with a "-"-sign). A lower (more negative: e.g. -50 V instead of -40 V) bias-voltage leads to smaller bias-current (and vice-versa), but this means that the larger absolute number (50 vs. 40) corresponds to the smaller bias current (and vice versa). All this is now connected to the one term "bias" in many not-so-professional publications. What does an author seek to express when he/she writes "turn up the bias"? Should it be more bias current (idle-current) i.e. plate-current (or even cathode-current!) when no input signal is present? Or more voltage fed to the grid via the bias pot? If the latter: more voltage in absolute numbers (i.e. go to from 40 V to 50 V, both voltages being negative), or higher voltage in terms of physics (i.e. go from -50 V to -40 V)? It's all rather complicated, and one person implies *this* while the other understands *that* – but only because (and if) unclear terminology is used. Therefore, let's talk about idle-current, or bias-current, or grid-bias-voltage or even bias-voltage (with a clear "-"-sign, and watching the polarity of our meter); but let's avoid "bias" without further specifics. That term is simply not precise enough.

♣ The corresponding scale (no lower boundary) includes the rating „has not a single clue whatsoever, at all".

In the following, the term "**bias-current**" is used to designate the plate-current flowing in *one* power tube when no input signal is present. Alternative terms for "bias-current" would be "quiescent plate-current" or "idle current". We will use the term "**bias-voltage**" to indicate the DC-voltage fed to the control-grid of the power tubes when no input signal is present. It is always a negative voltage. An alternative term for "bias-voltage" would be "grid-bias-voltage". This terminology will lead to e.g. the precise statement: "At a bias-voltage of -50 V, a bias-current of 38 mA flows." Only one possible pitfall remains: it needs to be clear that the bias-voltage is the voltage measured from grid to cathode. In case the cathode is connected to ground via a resistor (and possibly a capacitor in parallel), the grid/cathode voltage is not the same as the grid/ground-voltage. When taking a measurement, this is important. In the following, "bias-voltage" always indicates the grid/cathode-voltage.

Cold Biasing indicates that the bias-current is set to a relatively low value – corresponding to the "very negative" bias-voltage. In line with what has been said above, it is a bit problematic to use the terminology "small bias-voltage" because not everyone may understand that -50 V is smaller than -40 V. For a thermometer, the situation would be clear: -20 C° (or F°) is colder than -10 C° (or F°), and so -20 C° (or F°) is the colder/lower/smaller temperature. Despite a clear separation into topological and metrical scales, the Internet community has found its own interval scaling, and we may read the terminology "turning up the grid voltage" as a readjustment from -40 V to -50 V. Whether the voltage or the magnitude of the voltage is increased – it does make a difference. **Hot Biasing** is the other extreme: high bias-current, and a "less negative" bias-voltage. In other words, to safely avoid all doubts: the smaller the magnitude (i.e. the numeric value) of the (negative) bias-voltage, the hotter the tube is run. 10 mA/-60V is cold, 80 mA/-40 V is hot – just as an example! Because we will see in the following that these numbers are circuit-specific; one circuit's "cold biasing" may well be the other circuit's "relatively hot".

In **Fig. 10.5.21** we see both the characteristics of the individual tubes (dashed) and the overall characteristic generated by superposition. The center picture shows a "cold operating point" i.e. "**cold biasing**". With little change in the drive level (the voltage indicated on the abscissa), neither of the tubes feels animated towards much activity – both still are in blocking mode and the overall current remains small. Only for larger input voltages, the tubes start to move into the respective (alternating) conducting state and the current increases. The result is a saddle-shaped crossover-distortion. In the left-hand picture, the situation is different: the bias-currents in the operating point are higher, the overall (summed) characteristic retains its incline over a broad range and only curves clearly at the overdrive-limit.

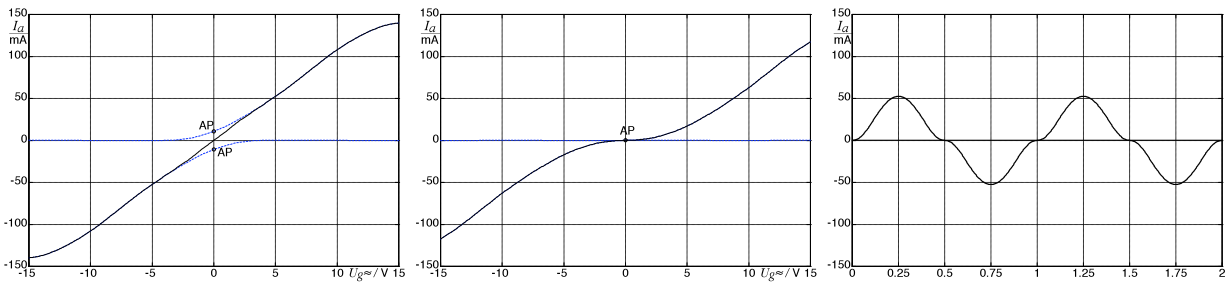


Fig. 10.5.21: Characteristics for two different settings of the bias-current. On the right, the distortion relating to the characteristic in the center picture is shown ("crossover-distortion").

The cold biasing shown in the center picture has two effects: undesirable non-linear distortion, and an expansion just as unwanted. In contrast to a compressor, an expander increases its gain with increasing signal level – not a stylistic device many guitarists welcome. Corresponding measurement data are shown in **Fig. 10.5.22**: in a Super-Reverb, the bias-current (I_a) of the power stage was varied between 10 mA and 53 mA. With a bias-current set to 10 mA, the output level increases by 27 dB as the input level rises by 10 dB – this is already a noticeable expansion. The right-hand picture shows the corresponding 3rd-order distortion level – again there are clear differences.

This figure is reserved for the print version of this book

This figure is reserved for the print version of this book

Fig. 10.5.22: Fender Super-Reverb (2x 6L6-GC), bias-voltage varied between –65V and –40V.
Left: output signal level vs. input signal level; right: distortion level vs. input signal level. **NFB disabled**.

In order to document the effects of the bias-current on the forward branch, the measurements for Fig. 10.5.22 were done with the negative feedback loop left open. With active NFB (**Fig. 10.5.23**) we see similar curves with a minimally weaker expansion and slightly lower distortion. For the "hot" operating modes the difference in the distortion is clearly visible; for the operation with low bias-current we need to consider that for equal input levels, the output levels differ considerably, after all.

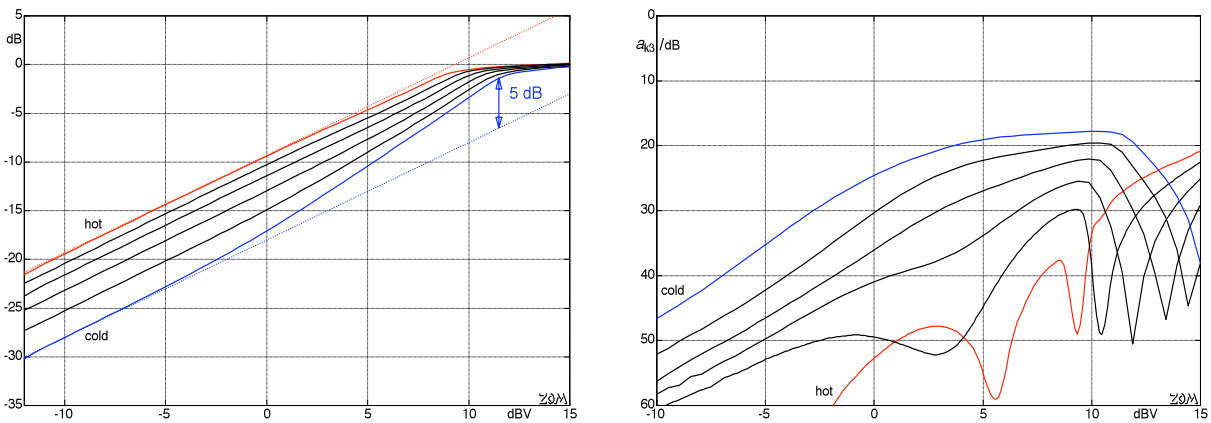


Fig. 10.5.23: Fender Super-Reverb (2x 6L6-GC), bias-voltage varied between –65V and –40V.
Left: output signal level vs. input signal level; right: distortion level vs. input signal level. **NFB enabled**.

Usually, the Super-Reverb will not be operated with a bias-current as small as shown via the blue curves. A larger bias-current (about 35 – 45 mA) would be normal. However, the bias-current should not be much larger, either, because the plate-loading would then possibly enter the critical range (see Fig. 10.5.26).

Variations of the bias-current in the power stage will change distortion, gain and dynamics, and also alter the internal impedance. We have already seen in Chapter 10.5.7 that the internal impedance of a tube is not constant but depends on the operating point. The internal impedance transformed by the output transformer therefore depends on the OP, too. With a high internal impedance of the power stage, the loudspeaker experiences less dampening, and resonances influence the transmission behavior more strongly. Moreover, since the speaker impedance rises towards high frequencies (Chapter 11), the high internal impedance results in a treble boost. **Fig. 10.5.24** shows measurement results for a Super-Reverb.

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Fig. 10.5.24: Super-Reverb. Left: magnitude of the output impedance; right: transfer-function of the power stage.
NFB active; amplifier loaded with loudspeakers (4x Jensen 4xPR-10). Different bias-currents.

Operating the amplifier with a small bias-current (cold biasing, I_a = 15 mA), the internal impedance of the amplifier (with active negative feedback) amounts to 30 Ω – this is relatively high compared to the load impedance. The resonance peak and the treble boost are more pronounced than for the "hot biasing" shown in red (I_a = 50 mA).

A different picture emerges for the Marshall power amplifier which features stronger negative feedback. The output is of significantly lower impedance, and the loudspeaker impedance maps onto the output voltage to a much lower extent (**Fig. 10.5.25**). For usual settings of the bias-current, the 16-Ω output of the JTM-45 is lower in impedance compared to the Super-Reverb by a *factor of five*! However, it would be wrong to conclude that the Marshall could/should be operated with a loudspeaker of smaller impedance (or the Fender with a loudspeaker of higher impedance): the optimum load-resistance is not directly derived from the internal impedance but from the limit data of the tubes.

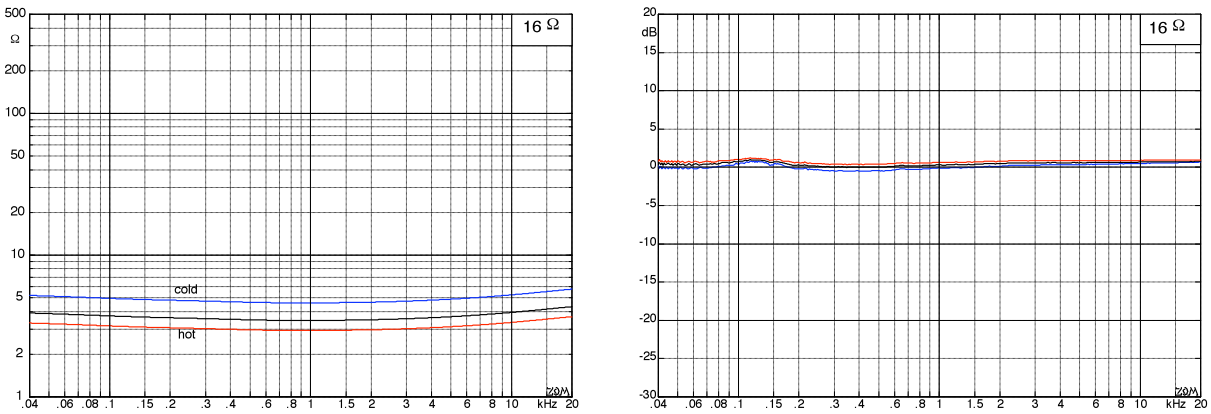

Fig. 10.5.25: JTM-45. Left: magnitude of the output impedance; right: transfer-function of the power stage.
NFB active; amplifier loaded with loudspeaker (Marshall 1960-AX). Different bias-currents.

The above pictures show that the transmission characteristics (and therefore also the sound) of a power stage depend on the setting of the bias-current. The latter also influences the **power dissipation**, and the following is dedicated to this issue. The less negative the bias-voltage is, the larger is the plate-current and the larger the power dissipation at the plate. We frequently read that the power dissipation at the plate (without input signal) should be 70% of the maximum permissible power dissipation. As an example: for the 6L6-GC, specified at 30 W, this would be 21 W (e.g. 47 mA at 450 V). It is not purposeful to search for the origin of the 70%-rule – that would be much too speculative. More conducive is to build an example explaining the strain that the tubes experience. **Fig. 10.5.26** shows three load lines of the 6L6-GC drawn into the output characteristic. For the upper line, we assume a bias–current of 47 mA, and for the lower line one of 33 mA. We calculate (for a plate-voltage of 450 V) a power dissipation in idle of 21 W (70%), and 15 W (50%), respectively. Although the dissipations in idle differ by as much as 40% (21 = 1.4 · 15), the maximum power for R_a = 1.3 kΩ varies by only 7%. Only changing the load-impedance from 1.3 kΩ to 1 kΩ brings larger differences in the strain on the plate. What about the mean values of these curves? They depend on the individual drive levels. The worst-case would be a square-shaped plate-current of an amplitude of 200 mA (1.3 kΩ); the determined instantaneous power would have to be halved because each power tube conducts only for one half-wave. For R_a = 1.3 kΩ, the maximum allowable power dissipation at the plate is not reached – it is, however, already slightly surpassed for R_a = 1.0 kΩ.

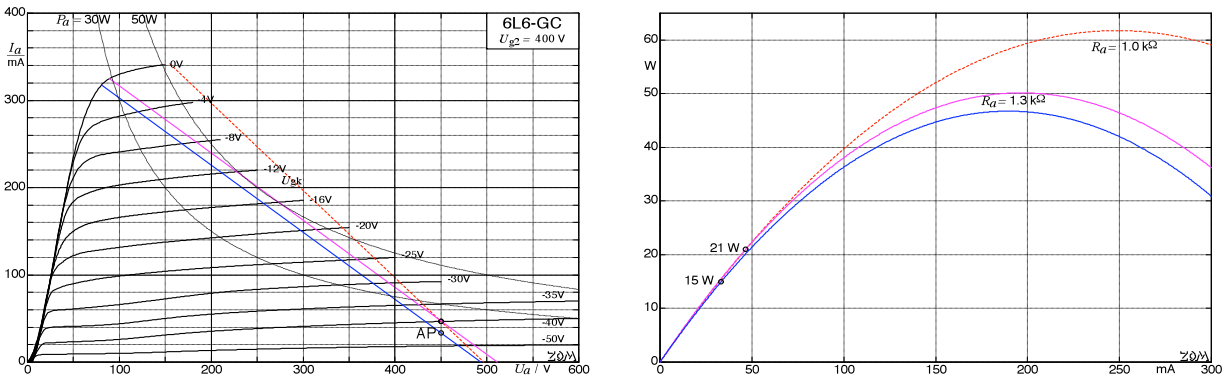

Fig. 10.5.26: Output characteristic of the 6L6-GC. AP = operating point without input signal.
Right: power dissipation at the plate dependent on the plate-current. Load impedance = 1.3 kΩ / 1.0 kΩ.

In conclusion: regarding the strain on the plate, the load-impedance is much more important than adjusting the bias-current to the second decimal. If the load impedance becomes too small, the plate will be overloaded. Of course, the type of drive- (or overdrive-) signal plays a role, as well – as does the voltage at the screen-grid … and as does the plate-voltage. Most everything that can change does change. Therefore there is no harm in calculating a load line once in a while – but only if we do not seek to adhere slavishly to the results. In Fig. 10.5.26, the operating point was assumed for 450 V. However, with the presence of a drive signal, the voltage delivered by the power supply does not remain constant but may easily change by as much as 50 V depending on the load… the strain on the plate will change correspondingly. Also, the load line will deviate even much more from the normally assumed straight line. No **loudspeaker** has a constant and purely ohmic impedance (Chapter 11). Rather, the loudspeaker impedance is complex (voltage and current are phase-shifted re. each other), and its magnitude can easily vary by a factor of 10 depending on frequency. Calculations using straight load lines are highly idealized models – nothing more but also nothing less. Reality is very different, in any case.

In reality, a guitar amplifier is neither driven by sine-tones nor is it loaded with a purely ohmic resistance. Neither its supply voltage nor the voltage at the screen-grid are constant. **Fig. 10.5.27** shows a first step in the direction of reality: these output characteristics were not determined on the test-bench but from a real guitar amplifier.

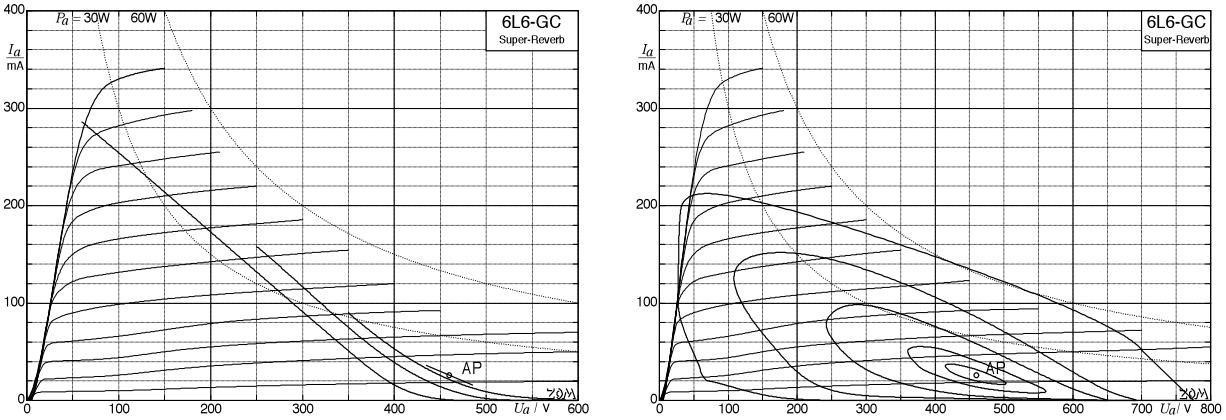

Fig. 10.5.27: Output characteristic; Super-Reverb with purely ohmic load (left), and with complex load (right).

For the measurements shown on the left, the amplifier was loaded with a purely ohmic 8–Ω–resistor. Small drive levels yield a small, straight line through the operating point. This line grows as the drive level increases, bends and shifts to the left. Consequently, even an ohmic load does not generally warrant assuming a straight line passing through the operating point. This is because on one hand the supply voltage drops, and the other hand the coupling capacitors are polarized due to the current flowing through the grids (Chapter 10.4). The curves on the right are for a complex loudspeaker-load (f = 3 kHz). For small input levels we see ellipses encompassing the operating point; large drive levels result in sharply bent curves that extend into the range of 30 W – which has been specified as limit. Since that value needs to be seen as short-term power average, this transgression does not generally indicate a thermal overload of the tubes (Chapter 10.5.9).

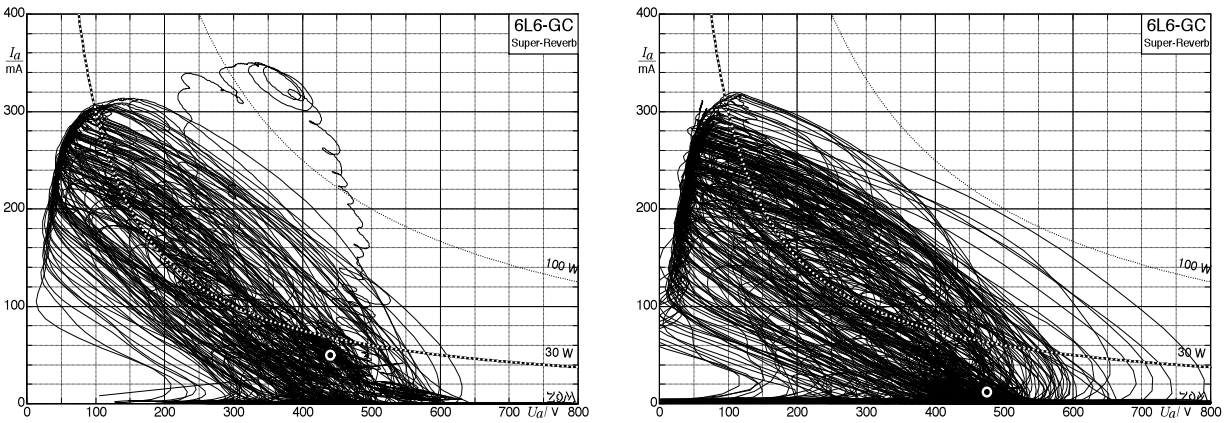

Fig. 10.5.28: Output characteristic, Super-Reverb with loudspeaker load using guitar tones (Stratocaster).

Even closer to reality are the curves depicted in **Fig. 10.5.28**: with a loudspeaker as load, the Super-Reverb was played with a guitar – and no sign of any straight load line at all remains! Rather, there is a myriad of highly different loops that only with great difficulty allow for any conclusion regarding the setting of the bias-current (white circle). Therefore, the load line is unsuitable to establish any connection between bias-current and power dissipation in the tubes – to do this, true measurements of the power dissipation are necessary.

In order to measure the power dissipation at the plate, the plate-voltage and the plate-current need to be recorded. Just multiplying the RMS-value of the plate-voltage with the RMS-value of the plate-current is not sufficient because that way we would merely determine the *apparent power* [20]! So: anybody connecting a volt-meter to the plate, and an ampere-meter serially into plate-connection, will indeed measure U_a and I_a, but the product of these values will only give information about the strain on the tube in a DC-situation. With an input signal present, however, AC results – and here we need to distinguish between effective power, reactive power and apparent power. It is the **effective power** (the product of plate-voltage and plate-current averaged over time) that heats up the plate. It is important to understand that it makes a difference whether the multiplication comes first (ahead of the averaging – correct for the present considerations) or averaging comes first (ahead of the multiplication – incorrect in the present case). Fig. 10.5.28 has impressively shown that for this 6L6-GC, the short-term dissipation at the plate exceeds 100 W – more than three times the value specified as a maximum. A tube needs to be able to take such a short-term overload if it is to be successfully deployed in a guitar amplifier. As we switch on the plate-current, the temperature of the plate begins to rise – thermal energy is supplied. At the same time, **thermal energy** is dissipated via radiation, and after some time a steady equilibrium, i.e. a constant plate temperature, is reached. If this temperature is too high (with the plate glowing brightly), the tube dies. If we do not wait until the equilibrium is in place, the temperature remains below the steady final value. Compare this to a car: stepping on the accelerator for only 2 seconds will not give you maximum speed. For a tube, though, 2 seconds would already be relatively long – in any case the typical short-term overload situations in a guitar amplifier will be of lesser durations.

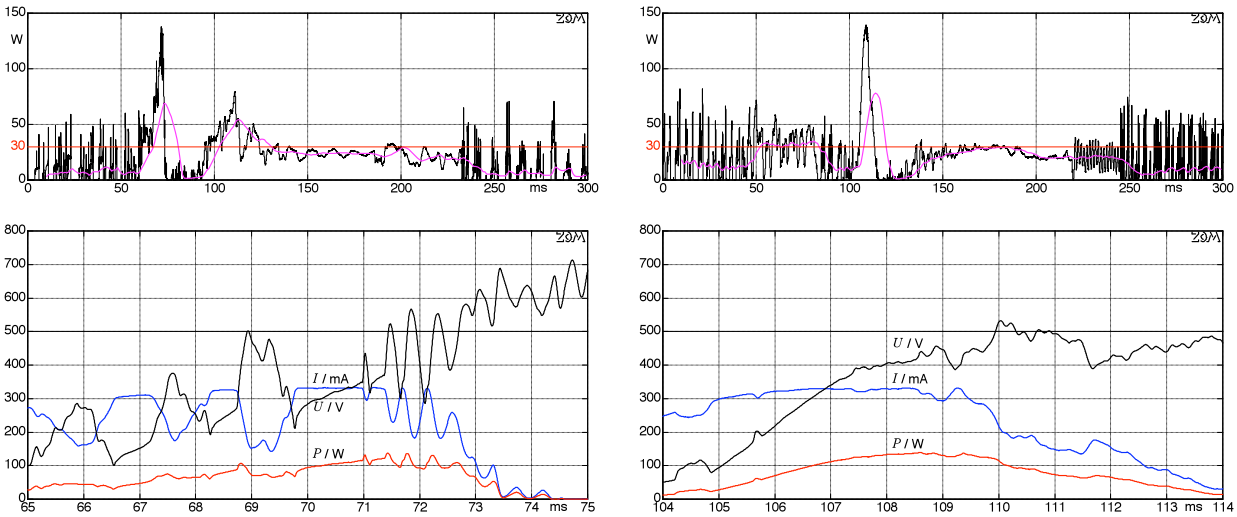

Fig. 10.5.29: Instantaneous power dissipation at the plate (black), sliding 10-ms-average (magenta).
The lower section shows excerpts from the progress of the plate-voltage, -current, and -power-dissipation.

Fig. 10.5.29 shows two examples for a loudspeaker-loaded Super-Reverb. The instantaneous values of the plate dissipation (sampled at 48 kHz) reach about 140 W. Averaging over 10 ms still gives values significantly in excess of the specified 30-W-limit. The lower line in the figure indicates that this strain happens in a frequency range unusually low for guitar tones: interpreting 10 ms as half a period, we get 50 Hz. This is not connected to the mains frequency but results from effects of overdrive-related recharging processes in phase-inverter (Chapter 10.4) and power stage. Since such short-term overloads can happen repeatedly, it is purposeful not to set the bias-current too high so that little effective power is fed to the tube in between such high-power phases. The question now remaining is: how high is a bias-current set “not too high?

So, at last: what is the correct setting of the bias-current (the "**bias setting**[♣]")? Unfortunately, there is no formula that will generally hold – the circuits, tubes, loudspeakers, and ways of playing that can occur are too diverse. Nevertheless here are some basic recommendations:

For **group 1** including laypersons i.e. persons without any education in electronics: whoever is not clearly aware of the reasons why in a 40W-amp voltages of over 800 V may occur, and who does not know how to protect him-/herself against the corresponding deadly dangers, must not open up an amplifier. Studying the manual of a multi-meter must not be understood as an education in electronics, and the same holds for confident handling of a screwdriver. Not everybody who removes an amp chassis from a cabinet instantly keels over dead – but this fact must not lead to the conclusion that this will never happen. If we are not allowed to open an amp, we can merely resort to measurements using a socket-adapter. The latter should be certified and re-checked regularly according to local regulations because it is subject of the same high voltages. Equipped this way, we now (more or less incorrectly) consider ourselves part of group 2.

Group 2 includes appropriately trained persons (e.g. certified electricians) who have simple measurement devices at their disposal. They should be in the position to adjust the bias-current without being in danger, should be able to recognize whether a power stage operates in true class-A mode (BIG exception), and then be able to find – using an oscilloscope – the middle of the load line. For an amplifier working in class-AB mode, the only helpful approach is a mixture of listening-tests and simple measurements of the power consumption in idle: if the amp already sounds good at 50% of the allowable plate-dissipation (e.g. 450 V, 33 mA for the 6L6-GC), you should just let it be. If your hearing (or the musician looking over your shoulder) demands more, you can run the thing a bit hotter – but at 70% most practitioners will raise an admonitory finger although there is no theoretical foundation for this limit. In any case, the power tubes need to be looked at in the dark to check whether, during any phase of extensive and multifaceted testing, grids and/or plate are visibly glowing. That this testing is not to be done with just a sine-generator and simple load-resistor should be – in view of the above – crystal clear by now.

Group 3 includes persons belonging to group 2 who have special instrumentation equipment in their arsenal, for example a current clamp that can measure with a resolution of 10 mA or better, and in the frequency range of 0 – 10 kHz. Seriously: 0 Hz – because the DC-components need to be measured, as well, and thus a frequency limit of 1 Hz is useless. Suitable would be e.g. the Tektronix AM305B/A6302, with the offset continuously monitored. Given such a current-measuring device and a high-voltage test-probe for the voltage measurement, you can then capture the factors determining the plate dissipation, digitize them in the calibrated front-end, store them in the computer and derive the actual, true loading of the tube. Once you've gotten that far, inevitably the question will arise whether today's tube manufacturers will actually still adhere to the tube data from the 1950s, and will warrant e.g. 800 hour MTBF for their products. The other immediate question is whether indeed every tube-wholesaler who allegedly cooperates in the development of "his" special tubes will expend such an effort.

In the case that "no" is the answer to these questions, we quickly move to become members of **group 0**. Here we join all those who have noticed that old Marshalls or Fenders did not even have any means to adjust the bias-current – but still did their job admirably. And so we change the tubes, if need be, and that's it.

[♣] too much of a bias never is a good thing – that seems to hold for all aspects in life.

10.5.9 Tube-strain and -aging

During regular operation, the power tubes of a guitar amplifier become very hot: 250 °C can be easily found at the glass container, and a much higher temperature within it. It is this heat that makes the utilized materials age, that will destroy coatings and consequently deteriorate the operational data. Of course, tubes may also break – but that is not normally the reason for their failure. Common is:

- The cathode coating evaporates from the cathode and deposits on other electrodes; overload accelerates this process.
- An insulating intermediate layer may form on the cathode between the carrier (e.g. nickel) and the emitting layer (mostly barium-oxide).
- Gases released from the electrode metals impair the vacuum.
- Mechanical vibrations bend or destroy electrodes.

In the MOV-datasheet for the GEC-KT-66 we find the lifetime specified at "minimum **8000 hours**". Consequently, an amateur playing 10 h per week at high volume would not need to be worried for 15 years, and even the pro (with 8 h per day of loud playing) would be able to enjoy that set of power tubes for 3 years. Author Helmuth Lemme [1995] quotes entirely different time-spans: he recommends to tentatively change the power tubes after **100 hours**. Edward van Halen apparently acts even more rigorously: allegedly, he has (or had) all power tubes replaced after every gig. Checking the MOV-definition for "end of tube life" more closely, the 8000-hour-euphoria is brought down a peg or two: we find that the output power has gone down to 50% i.e. as long as a 100-W-amp will yield 50 W, the tubes are deemed o.k. A guitar player will hardly be content to work with half the specified power, though, and therefore the MOV-definition is lacking in practical relevance. MOV is silent about the characteristic of the drop in power, we only find that reducing the strain on the tube by 40% will extend the lifetime by 25%. That does not help us to draw any conclusion about the lifetime under overload conditions. The latter often appear in guitar amplifiers; operating outside of the specified ranges is often the case. The recommendation to replace the output tubes after 100 h to check is therefore not entirely without merit. We will happily avoid discussing the occasionally found (wacky) idea that tubes should be "run in" for about 100 hours before they sound right; rather we will opt to clarify the question which operational state puts the highest strain on the tubes.

Just switching on an amplifier can be detrimental: subjecting the still cold tube to the full plate-voltage may cause parts of the cathode-coating to detach. Here, the good old **rectifier tube** did have an advantage: only once it had heated up, the full supply voltage was available, and by that time, the other tubes generally were at operating temperature, too. On the other hand, just keeping the filaments powered up without any current through the cathode should be avoided, as well, because it supports the build-up of the impeding intermediate layer (exceptions are the so-called long-life tubes). Whether an amp should, during breaks in playing, continue to run with the plate-voltage switched on, or switched off (i.e. on standby), or should be powered down completely is discussed controversially. Complete shut-down does reduce the "hours in action", but it brings numerous strong temperature fluctuations that also reduce the tube-lifespan – better leave the power on, then. Regarding the use of the stand-by mode, there are only assumptions: advantages and disadvantages are more or less in balance. In amplifiers with a high-bias current, the stand-by mode <u>can</u> be purposeful because in these amps the tubes are under the highest strain without an input signal. An example would be the VOX AC-30: in idle-mode, the maximum strain on the plate of the EL84 is usually already exceeded – but it is exactly this amp that does not have a stand-by switch.

What does put the strain onto the power tubes? The heating (or even over-heating) of the electrodes! Without a technical-education background, one could assume that the amp being operated at full drive would bring the power tubes to their strain-limit; overdrive would then cause **overload**. That is, however, not correct per se: relevant for the power dissipation at the plate is the product of plate-current and plate voltage. For example: in the idle-state, there are 450 V between plate and cathode♣, and the bias-current is 40 mA. The plate dissipation then measures 18 W. If the tube can take 30 W, this strain in idle is not critical. As a drive signal is fed to the amp, both U_a and I_a change. The change, however, in opposite directions: as I_a rises, U_a falls. At the drive-limit, the plate dissipation would even converge to zero (at least for idealized conditions): either the tube conducts; in that case there is a plate-current but the voltage drop across the tube is zero. Or the tube is in blocking state; now there is a high voltage across the tube but the current is zero. No power tube is that ideal, though: at maximum plate-current, we find a voltage of about 50 V between plate and cathode, or even more. Still, even with a plate-current of I_a = 0.3 A, this would imply merely 15 W i.e. totally uncritical. The danger lurks in the intermediate range at around half of the drive-level range: 225V · 0.15A = 34 W. With a 30-W-tube, we would be already outside of the specified **strain limit**. The latter needs to be seen as an average value, though – the tube is not operated statically in this state for the duration since the input signal changes all the time. Here, we find strain calculations that determine the plate dissipation for sinusoidal drive signals. This is not entirely unreasonable, but not typical, either: the signal delivered by an electric guitar is not sinusoidal. Alternative calculation methods assume a square drive signal and result in a 23% higher strain on the plate – however, the guitar signal is not always of a square shape, either. In any case: even assuming worst-case as continuous operational state, the plate will take on the thermal strain quite nicely – at least as long as the load impedance fits.

But then there's the **screen grid**! Contrary to the situation at the plate, the voltage at the screen grid decreases insignificantly in the presence of a drive signal, therefore it remains an ideal landing spot for the electrons when the plate-voltage is small. Consequently, current and dissipation in the screen grid increase as the plate-voltage drops. Since the maximum allowable dissipation in the screen grid is smaller than the maximum allowable plate-dissipation, the screen grid can easily be made to **glow**. Moreover, it is much more difficult to observe this compared to a glowing plate because the screen grid is surrounded by the plate. Here is a numeric example: the 6L6-GC is specified with $P_{a,max}$ at 30 W and $P_{g2,max}$ at 5 W. At full drive level (U_B = 400 V, U_{g2} = 350 V), the plate experiences a strain of 19 W, and the screen grid of no less than 15 W! Such an extreme overload must only be present for short periods unless we want to run the risk of the screen-grid wires melting and the tube dying. This is why, during the circuit design, the screen-grid dissipation is measured, as well, and measures to limit it are taken if necessary. The tried-and-tested method here is to introduce a **screen-grid resistor** in series with the grid. In fact, this resistor has two functions: to suppress RF-oscillations, and (given the appropriate resistance) to decrease the voltage at the screen grid for high currents through the screen grid. Some old **Marshall-** and **Fender-amplifiers** lack any screen-grid resistor, and the lifetime of the tubes can become extremely short – in particular if the **EL34** is deployed. This tube is a true pentode and the current through the screen grid can easily reach values 2 – 3 times as high compared to the 6L6-GC. If a screen-grid resistor is present, it often is given a value of 1 kΩ. This resistance is recommended as adequate to avoid RF-oscillations, but the screen-grid dissipation may not be reduced enough. You could raise the resistance to 5 kΩ, but that may have effects on the output power and the sound (the latter of course being a matter of taste). Sound or safety – you choose.

♣ That would be for class-AB operation; for class-A operation, the strain on the tube is highest in idle.

The family of output characteristics gives us a clear overview regarding the relation between current and voltages (**Fig. 10.5.30)**. Changing the drive levels shifts the operating point along the load line and changes the power dissipation at the plate. The fact that the load line traverses the power-hyperbola, and that a peak value of 44 W results, is not critical because the operating point advances into the range of such high dissipation only for one half-wave. In the worst case (a square wave-shape), the thermal strain on the tube is 22 W, which is still clearly below the 30-W-limit.

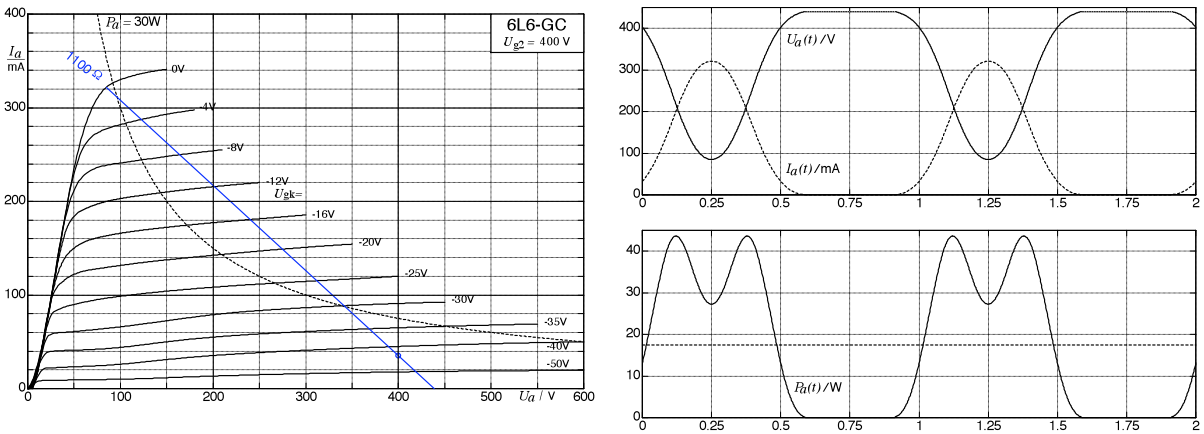

Fig. 10.5.30: family of output characteristics of the 6L6-GC for a load impedance of 1.1 kΩ; time functions for a sinusoidal drive-signal. In a push-pull output-stage including an output transformer, the plate-voltage would not be limited to 440 V (as indicated here) but rise to 700 V (Chapter. 10.5.3).

Fig. 10.5.30 is for a purely ohmic **load-impedance** of 1.1 kΩ. If this value changes, the gradient of the load line changes, as well, and with it the strain on the tubes. Increasing the load-impedance (smaller load-line gradient) reduces both plate-current and dissipation at the plate. Decreasing the load-impedance increases the strain on the plate: a power stage specified for 8 Ω should therefore not be operated at 4 Ω for extended periods of time.

We find entirely different functions for the **current through the screen grid** (**Fig. 10.5.31**). Given a constant voltage at the screen grid, the maximum power dissipation at the screen grid rises to 125 W – if at all, this is only allowable for impulse-operation: according to the datasheet, 8 W should not be exceeded. Even with 1.5 kΩ connected between voltage source (again 350 V) and screen grid, the allowable screen-grid-dissipation is, at 20 W, considerably exceeded. The same holds for the 6L6-GC (**Fig. 10.5.32**). On the other hand, the question remains whether this state can actually happen during real operation?

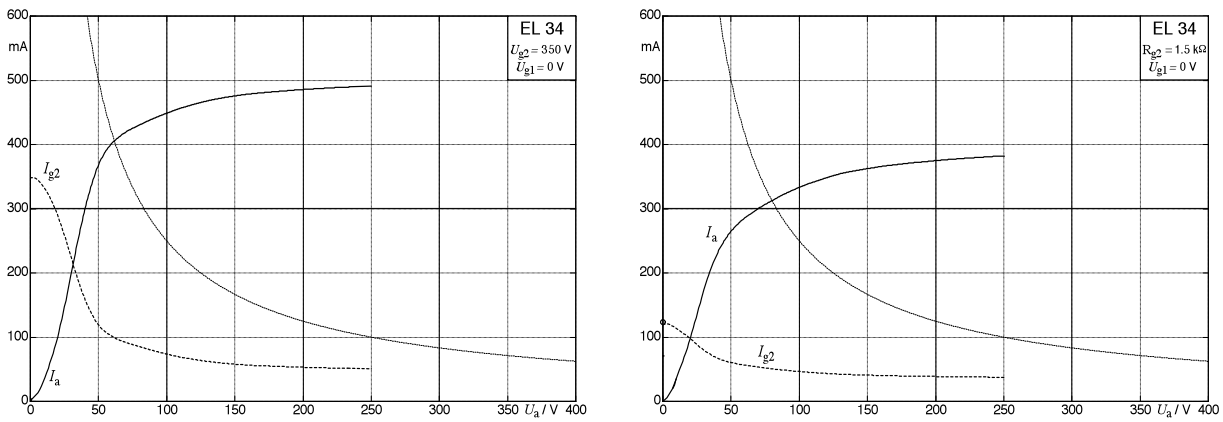

Fig. 10.5.31: Plate- and screen-grid-current dependent on the plate-voltage.

Fig. 10.5.32 depicts the average power dissipation at plate and screen grid, dependent on the drive levels (square-shaped drive signal). The highest plate dissipation P_a appears at $U_{g1,\max}$ = -14 V; for R_a = 1100 Ω the average dissipation is P_a = 22 W, which is safely below the strain limit. A higher load-impedance (1500 Ω) reduces the strain on the plate, and we generally find that **a higher-impedance loads relieve the plate**. The maximum power dissipation at the screen grid P_{g2} happens at $U_{g1,\max}$ = 0 i.e. at a fully overdriven power tube. For a load impedance of 1100 Ω, P_{g2} is 8 W, and for a 1500-Ω-load we already see 14.5 W – three times the allowable power dissipation at the screen grid. For overdrive conditions we therefore find: **higher-impedance loads put more strain on the screen grid**.

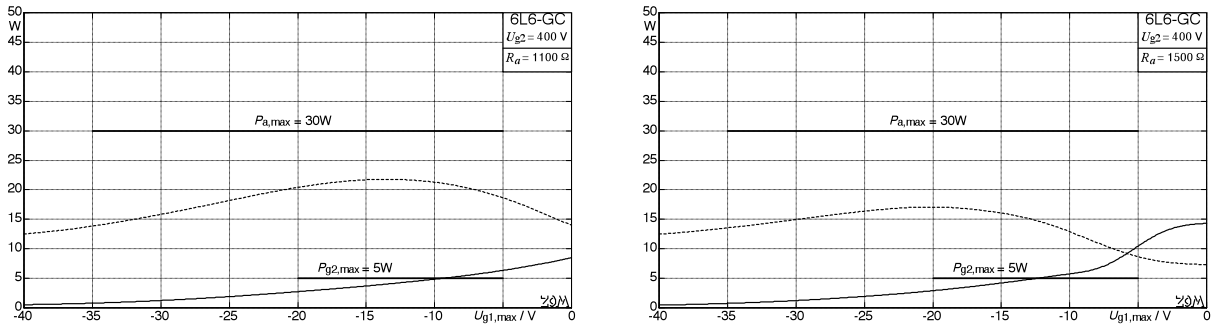

Fig. 10.5.32: Symmetric square-shaped drive around the operating point of 45 V / 30 mA, power dissipation at plate and screen grid (averaged over time). No screen-grid resistor

Consequently, a tube amplifier can in no way be seen as totally immune against a **load-mismatch**. The 8-Ω-output should indeed be connected to an 8-Ω-loudspeaker! Too small a load-impedance (e.g. two 8-Ω-speaker-boxes in parallel = 4 Ω) would increase the strain on the plate – although this would not lead to immediate failure. On one hand, there is some overhead here, and on the other hand, the higher stain will reduce the supply voltage (depending on the power supply circuit) such that the strain is not quite as high. Too high a load-impedance (e.g. a 16–Ω-speaker connected to an 8–Ω-output) will increase the strain on the screen-grid – in particular if the power stage is operated often under overdrive-conditions. In this case, there is little reserve, as everybody measuring the power stage (connected to a loudspeaker) with a sweep may notice right away: every speaker will turn high-impedance at high frequencies, irrespective of its nominal impedance (Chapter 11). With such a high-impedance load, even a single measurement can lead to immediate failure of the power tubes. Looking at the datasheet and considering a load-impedance of double the optimum value, we find a static power dissipation at the screen grid of 40 W: P_{g2} = 0.1 A · 400 V (at U_{g1} = 0 V). The tube is in active state during only one half-wave such that on average 20 W remain, but that is still much too high compared to the allowable max. 5 Watt. Also, the speaker impedance may not merely double – a 10-fold increase is possible, as well.

In view of all this, the impression manifests itself that the classic power-stage circuits were not developed for Hardrock but for radio programs. And indeed, how could it be otherwise: 70 years ago, the place of action for a 6L6 was a radio receiver in most cases (or maybe an amplifier in a cinema, at most). Guitar amplifiers were few and far between. Even if such a tube found its way into such an exotic job site, it still lived a relatively tranquil life – the rocker "turning everything to 10" was only just about to be born. He (or she) arrived only later, but then used amplifiers that were – with unbelievable tenacity (or ignorance) – built as if reproducing moderate dance music without distortion were the only way of life. Marshall model 1987 amps of certain vintage include two EL34 but no screen-grid resistors. Lucky are those who can after each gig afford a new set of tubes (at 50 to 100 Euro).

We can insinuate that the developers of the early "original circuits" strove to avoid overloading the tubes too much at least in idle mode. No guitar player will buy an amp for its idle-mode qualities, though, and will "turn it up" at some point. Several parameters determine how the strain on the tubes will then change: the drive levels, the loudspeaker impedance, the specific tubes used, and the circuit variant. To start with the latter aspect: most amplifier circuits went through a number of stages of advancement, often driven by the demand for more power. When the Fender Twin entered the market in 1952, it was specified at 18 W. Only 3 years later, this power output had grown to 30 W, then 60 W and 80 W, and finally to 100 W (and beyond). The 5C8-Twin feeds 370 V to the plates – in the AC568 this is 470 V. A VOX AC-30 expects its EL84 to each put up with 14 W in idle, if it is operated with the rectifier tube customary back in the early days. Exchanging the rectifier tube for silicon diodes (a design development in later AC-30's) pushes the power dissipation at the plates to 17 W each. Marshall amps may or may not sport screen-grid resistors (25 Ω, or 470 Ω, or 1 kΩ), and the tube complement could include KT-66, 6L6-GC, EL-34, KT-88 or 6550. Therefore even if amps look similar, the strain on the tubes may differ significantly.

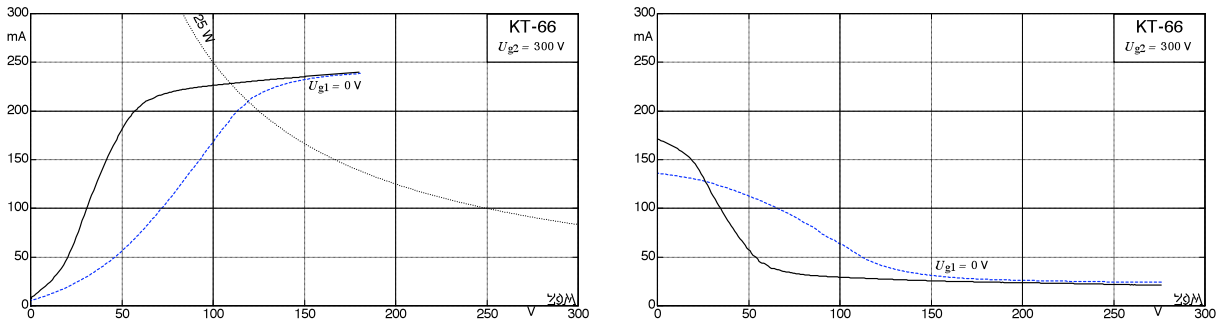

Fig. 10.5.33: Characteristics of two tubes both sold as KT-66; plate-current (left), screen-grid current (right).

Occasionally, even the tube designations may be just as unspecific: the measurements shown in **Fig. 10.5.33** were taken from two fresh KT-66. They show significant differences both in the achievable maximum output power of the amplifier, and in the strain on the tubes, despite the fact that allegedly this is the same type of tube. To be sure, most tubes sold today will be roughly in the ballpark of the datasheet specifications; however, the amount of deliberately sold "selected" defectives unfortunately is not petty – to put it mildly.

It requires no emphasizing that the strain on the tubes depends on the speaker impedance, too: loudspeaker impedances are complex and strongly dependent on frequency (Chapter 11.2), and therefore the load line deforms into an ellipse in real operation, rendering immaterial all calculations of power dissipation for nominal load. In addition, the input signals almost never correspond to the textbook-sinusoidal shape: power stages in guitar amps are often overdriven. Not always, admittedly – but even with a seemingly "clean" sound, the string attack can drive the power stage into short-term limiting. More than a few guitarists appreciate their tube amps especially because of the power-stage limiting which is not easily imitated via small effects boxes (in contrast to the pre-amplifier distortion). Under overdrive conditions, a current flows through the control grid polarizing the coupling capacitors such that the operating point drifts back and forth depending on the drive signal – not an effect that is typically discussed in circuit-design textbooks. The latter will give you guidance to push the THD below 1%, and will exemplarily calculate the whole HiFi power stage. Continuous overdrive is not a topic here, nor was it in 1940 for the amp-forefathers. Only today, in everyday stage-life, it very much is.

In order not to succumb to the temptation to join in and explain the operating characteristics of a guitar amp only in the linear range, let's turn to 10 now – proud 'n' loud. The candidate for our measurements is a Fender Super-Reverb, its two 6L6-GC generating an output power of approximately 40 W – or more if we go into overdrive. Which we do: **Fig. 10.5.34**.

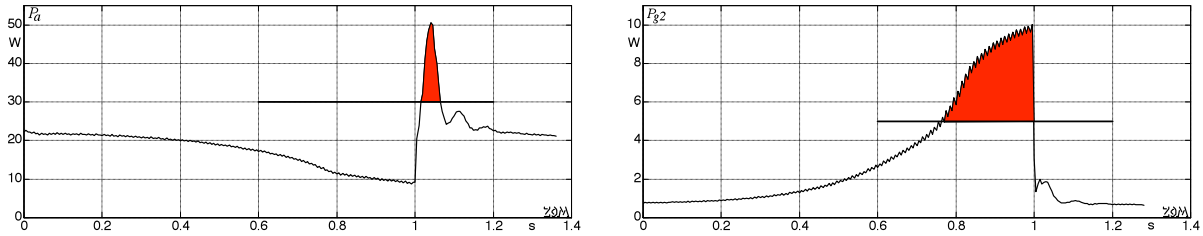

Fig. 10.5.34: Average power dissipation at plate and screen grid of a 6L6-GC, Super-Reverb. I_{Bias} = 50 mA.
Sinusoidal 1-kHz-tone with a level increasing by 20 dB from 0 – 1 s; switch-off at t = 1 s. Measurements taken with 8 Ω load (purely ohmic) for plate dissipation, and 16 Ω load (purely ohmic) for screen-grid dissipation.

The drive signal is a sine-tone with a level increasing by 20 dB from the time "0" to the time "1 s"; at t = 1 s we switch it off. During the last quarter of the measurement, the power stage is overdriven, which does not harm the **plate** at all: its strain decreases to about 10 W with increasing drive level. After (!) switching-off, however, there is a short-term plate dissipation of 50 W due to the settling of the polarization of the coupling capacitors. The tube does not die right away because this overload happens only for a short time. If such short-term overload conditions repeat themselves quickly one after the other, they could, however, pose a problem, after all. Also, 50 W is not really the end of the line: corresponding measurements with a real loudspeaker as a load resulted in more than 100 W!

We see a rather different behavior for the power dissipation in the screen grid: it grows with increasing drive level, and approximately at the point where overdrive occurs it crosses over beyond the maximum value of 5 W. Therefore: as soon as the power stage is overdriven, the **screen grid** enters the danger-zone. If we could address one of the design-forefathers with this problem, the answer would probably be: "you don't overdrive the power stage!" Yeah you do, these days. The argument that the Super-Reverb is an amp for rhythm-guitar that should be played "clean" could easily be countered in that the power-stage design for this amp corresponds to the Fender-standard of the 1960's – the power stage of the Bassman (as just one example) is in no way more less prone to be overdriven. This was all by-the-book design. At the time.

Overdrive is the joint cause for putting excessive strain on plate and screen grid; the exact effective mechanisms are specific to the respective electrode. Normally, the plate-voltage in a power tube decreases with increasing plate-current; at full drive level (U_{g1} = 0), the plate-voltage will be minimal. At this point, however, the plate becomes rather unattractive as a landing-site for the electrons (emitted by the cathode). The electrons are much more attracted to the screen grid that remains at a high potential (high voltage), and they land (at full drive level) on the thin screen-grid wires. The latter promptly heat up under this bombardment and start to glow. Even datasheets do not shy away from specifying a power dissipation of 100 W or more for the screen grid (at U_a = 0) – and at the same time they will give a maximum strain of 5 W. This is not a contradiction, because for short-term strain (impulses), the dissipation limit is higher. How high is not specified, unfortunately. At the plate, entirely different processes are significant: as long as the load-impedance of the power stage is not too low, the plate does not run into danger even under dramatic overdrive conditions. However, the coupling capacitors will vary their average DC-voltage during periods of overdrive, and during the following balancing processes, danger looms, after all.

The source for the balancing processes just mentioned is the phase-inverter (Chapter 10.4). The two signals generated by the phase-inverter are equal in magnitude and opposite in phase only for moderate drive levels. Strong drive levels shift the operating points in the phase-inverter, and the coupling capacitors change their average DC-voltage. As the input signal is switched off, the coupling-capacitors potentials return to their quiescent state within 2 s♣ – it is here that the peaks in the power-tube-specific strain occur.

Another problem may present itself at the fringes of the transmission range: at very high and very low frequencies, the two signals from the phase-inverter do not maintain exact opposite phases. The output transformer can generate its impedance-transforming magnetic field only in an efficient manner if the plate-currents support each other. If both power tubes conduct at the same time, the transformer has the effect of a bifilarly wound coil – with the effect that the inductance goes to zero, and merely the copper resistance of the primary winding remains as load impedance for the plate (e.g. 50 Ω): the plate will possibly be overloaded.

In **Fig. 10.5.35** we see once more the family of output characteristic of the 6L6-GC, with an I_a/U_a-characteristic measured for a JJ-6L6-GC. For a load impedance of 1200 Ω, the "knee" of the curve is almost exactly met; the strain on the screen grid at this point is 350 V · 46 mA = 16 W. As we increase the load impedance to 6000 Ω, the strain on the screen grid grows to 54 W. Assuming that this maximum strain occurs only during one half-wave, we may half that value – but the remaining 27 W still overshoot the allowable maximum value considerably. The approximate load-impedance for the specified match is 1200 Ω, i.e. 8 Ω for the Super-Reverb[2]. However, loudspeaker measurements show that the magnitude of the speaker-impedance will be larger that than the specified impedance both at high frequencies and at the speaker resonance. A primary load of 6000 Ω corresponds to a secondary load impedance of 40 Ω – this can easily be achieved with a loudspeaker. A guitar will not normally generate continuous tones at 15 kHz, but sustaining notes in the range of the speaker resonance are possible – and may be dangerous.

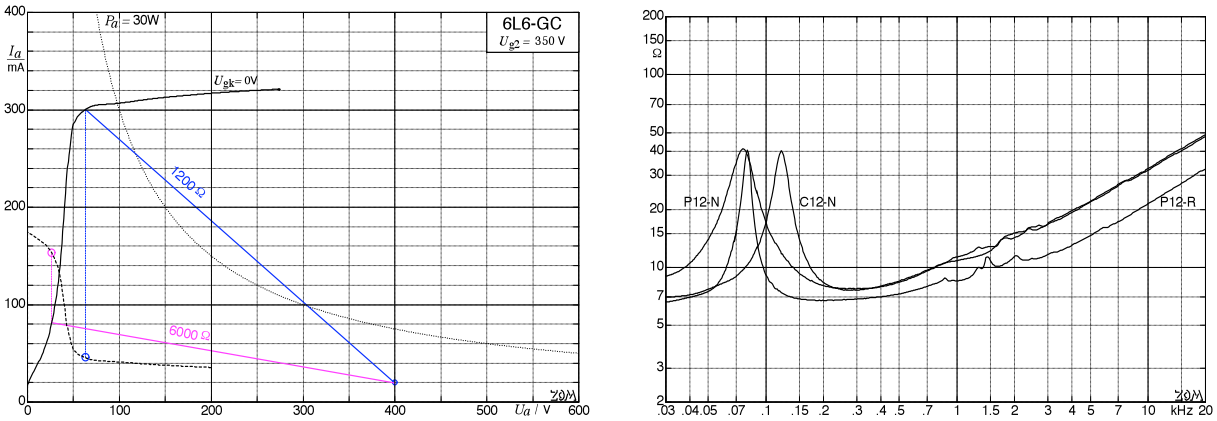

Fig. 10.5.35: output characteristic of a 6L6-GC (JJ) for two different loads on the plate (left), loudspeaker impedances (Jensen, right), plate-current (——) and screen-grid-current (---) for U_{g1} = 0.

In order to remain datasheet-compliant, the screen-grid resistor in Fig. 10.5.35 is assumed to be 0; in return, the screen-grid-voltage is only set to 350 V. A Fender-typical resistor-value would be 470 Ω, connected to 400 – 450 V. As a first-order approximation, we find similar strains on the screen grid; in the detail, there are differences that however cannot be calculated to the last watt.

♣ In theory, this asymptotic recharging takes an infinite time; 2 s should be seen as a specific guidance value.

[2] As already mentioned, this specific Super-Reverb carried a transformer with both 2-Ω and 8-Ω-outputs.

Often, the screen-grid resistors are connected to the first or second filter capacitors in the power supply. Without any signal at the amplifier input, these electrolytic capacitors charge to 400 – 450 V; with a strong input signal this voltage drops a bit (sags). How much the voltage drops depends on the load impedance and the internal impedance of the power supply. The sagging effect will be stronger for power supplies with a rectifier tube, and weaker for ones with silicon diodes. A power supply with rectifier tube and small caps (10 µF) will go easy on the screen grid while silicon diodes and 100 µF represent a challenge. Unfortunately, the datasheets for usual power tubes do not reveal any thermal time-constants of the screen grids, and therefore any calculation of the impulse-strain will remain speculative. Only with the triumph (?) of power transistors, impulse-diagrams enter the datasheets. The 2N3055 (aka BD-130), for example, is specified with 100 W continuous power dissipation at 20°C of the casing, with 320 W for 1 ms, and even with no less than 900 W for 30 µs. For the 6L6-GC, we find 5 W as a limit-value for the screen grid power dissipation, without any further details. In the semiconductor area, there is at least a rough guideline (in case you want to avoid much calculation) that the lifespan doubles if the operating temperature is decreased by 10°C. Given tubes, we need to rely on completely flaky speculations. How long did that Mullard survive 100% overload at the screen grid, and how does its modern Chinese remake fare? The remake that raises the suspicion that it's mostly the cosmetics that is important (it's got the brown base!). Caution, though! With such prejudice, you may well do very wrong by those sino-factures. Not everything that originates in China is bad – just as is the case for any other country. As he developed the 5881, was the Tungsol-R&D-guy in the US really interested in how strongly the screen grid would be overloaded in guitar amplifiers, and was that tube therefore marketed as the "better 6L6"? As late as 1962, the Tungsol datasheet specified: "*Maximum Grid #2 Dissipation: 3 Watts*". That's not really a lot, either, isn't it?

These days, acquiring a hand-wired boutique amp will easily set you back 4000 or 5000 Euro. That's without speaker, of course. Maybe the manufacturer boasts using only output transformers with original insulation-paper (with worse breakdown rating) and slightly rusted transformer sheets – to get the 'brown' sound? Maybe he will put a 1-A-fuse in the mains line (just as in the original) without realizing that converting from 110V to 220V the value of the fuse should be halved? It all has to be *original*- that's the main thing. Or, the focus is on using the same circuit that made the Bassman (or the Deluxe, the Twin, the JTM – you name it) famous. Including all the grid-destroying characteristics of these old amps. Amazing how the oldest cows are the most holy ones. Maybe it was the CBS-engineers who, by introducing protective circuits, discredited just these circuits. A guitar amp sounds best if it gobbles up a set of power tubes each evening – it's a cast-iron credo.

Rating Systems [Langford-Smith & RCA-Receiving-Tube-Manual]**:**
The **absolute maximum system** originated in the early days of valve development and was based on the voltage characteristics of battery supplies. Battery voltages could fall below their nominal values but seldom appreciably exceeded them, so that valve maximum ratings set on the basis of specified battery voltages were absolute maximum ratings that should not be exceeded under any condition of operation.
The **design center system** was adopted in the U.S.A. by the Radio Manufacturers Association in 1939 for the rating of receiving valves and since then has become the standard system for rating most receiver types of American design. Under the design center system, ratings are based on the normal voltage variations which are representative of those experienced with […] power lines. Design center ratings should not be exceeded under normal operation. These ratings allow for normal variations in both tube characteristics and operating conditions.
The **design maximum system** was adopted for receiving tubes in 1957. Design maximum ratings should not be exceeded under any condition of operation. These ratings allow for normal variations in tube characteristics, but do not provide for variations in operating conditions.

10.5.10 How does the 6L6 sound?

"6L6 = silky, clear treble combined with a well-defined, deep bass contingent ", states the advertisement. Or: *"EL-34 = delicate treble and well-defined bass and midrange."* Or: *"The 6L6 have more of a midrange tone"*. Or: *"The KT-90 will give a little more bass and treble, as compared to a EL-34."* On the other hand, we read in circuit-design textbooks that amplifier-tubes transmit their signal from 0 Hz up into ranges not measured in kHz but MHz. Osram, for example, recommends the KT-66 for radio-frequency amplifiers *"for frequencies up to 30 Mc/s"*. Much depends on the circuit the tube cooperates with, after all. In a first step, the parameters of the tube need to be taken as frequency-independent. That does not mean, however, that the whole power stage has no dependency on frequency. Also, changing tubes may well result in a different frequency response. That, however, does not allow for the conclusion that a special tube generally delivers more (or less) treble.

It is possible to investigate the behavior of a tube power separately for linear and non-linear operation, even though that is not entirely unproblematic for a guitar amplifier deliberately pushed into overdrive. But then, after all, there are guitar players who seek a sound as undistorted as possible. Also, if we focus on a system that distorts both linearly and non-linearly, we run into problems to describe it clearly (since no transfer function can be defined for such a system, amongst other reasons). In the approximately linear range, several components determine the (magnitude-) frequency response of the power stage: coupling capacitors (in combination with their load resistors), output transformer, and loudspeaker. For a guitar amplifier, we may not establish *one* frequency response of the power stage and *one* frequency response for the loudspeaker, and hope that these two diagrams would describe the overall system. For a loudspeaker fed from a stiff voltage source, we could define *one* frequency response (on axis), and for the speaker fed from a stiff current source, too – but these would be different frequency responses. For an internal impedance of 10 Ω of the amplifier, we would obtain yet another frequency response of the loudspeaker. For the power stage, the situation is similar: for an 8-Ω-load, the frequency response is different compared to a 16-Ω-load, and considering the loudspeaker loading results in different curves yet again.

One criterion in which power tubes may differ is their internal impedance. In pentodes (that are operated as such!) it is typically rather high: think 30 kΩ or so – but do keep in mind that this depends on the operating point i.e. on the bias current (Chapter 10.5.7 and 10.5.8). As the internal impedance of the tubes changes, so does the internal impedance of the amplifier. However, **negative feedback** (NFB) enters the stage at this point. For power stages with strong NFB (Chapter 10.5.6), a change in the internal impedance of the tubes has little effect on the internal impedance of the amplifier, but for power stages without NFB, these effects are considerable. Consequently, we may not conclude that characteristics found in an amp without NFB are also found in an amp that includes NFB. The higher the impedance of the amplifier output, the more resonances and treble range are emphasized. This, however, does not justify the purchase of expensive tubes (even though that may be suggested in ads): inexpensive components allow for varying the frequency response of the power stage within a broad range, as long as the linear characteristics are the issue. It may well be that a replacement tube has less gain that its predecessor – that can easily be compensated for by turning up the volume. If the power stage includes NFB, changes in the loop gain could influence the frequency response (Chapter 10.5.6), but it is easy to get a handle on this, as well – and with simple means. In linear operation, any frequency response can be achieved with any power tube; that is standard engineering-knowledge.

Things get more difficult in the non-linear range. Both phase-inverter and power stage may be overdriven. If high-gain power tubes are used, we can assume that these will distort first but if they have only a small gain, the distortion of the phase-inverter comes to the fore. Talking about power-amp distortion, we must therefore also keep the **phase-inverter** in mind. Your typical cathodyne circuit (Chapter 10.4.2) can generate up to about 40 V of voltage-amplitude, which is not enough if the grid-cathode-voltage of the power tubes is set to - 50 V. The paraphase- and long-tail-circuits, however, can easily drive the power tubes to their limit (and beyond) even for -60 V. It is necessary to always consider the power stage as a whole.

One possibility to describe the non-linear behavior of tubes is offered via the family of output characteristics, an alternative to this would be the transfer characteristic $I_a(U_g)$. All power tubes have frequency-independent amplification parameters throughout the audio range – their maximum currents are, however, dependent on the load. Since the load impedance (i.e. the loudspeaker) is frequency-dependent, there is also a tube-specific dependency of the maximum obtainable power output. **Fig. 10.5.36** shows the output characteristics of two power tubes. For a load impedance of 2200 Ω, tube A will give more maximum power, while tube B will have a higher output power at 550 Ω. At 1100 Ω, both offer the same maximum power. In this example, the maximum power is impedance-dependent, and since the impedance is frequency-dependent, so is the maximum power. The right-hand picture indicates that for higher load impedances a current-saturation happens already at U_{g1} = -8 V – the transfer characteristic turns into the horizontal.

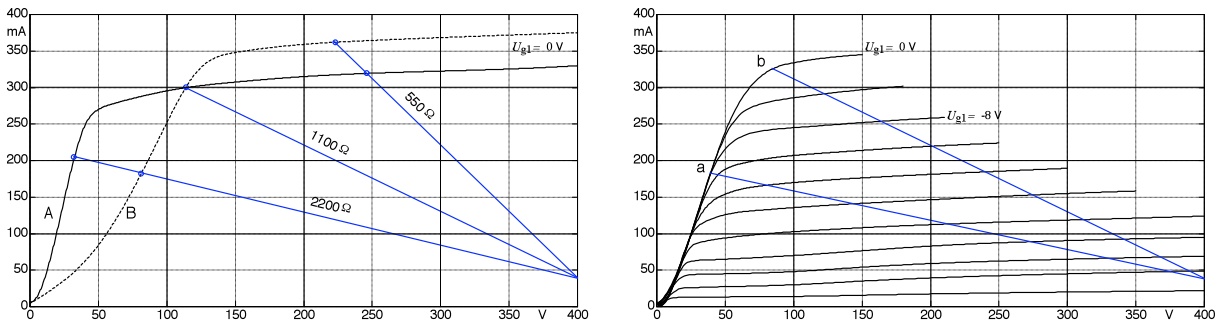

Fig. 10.5.36: Family of output characteristics: plate-current vs. plate-voltage.

Already these initial considerations show that exchanging the power tubes can change both the linear and the non-linear behavior of the power stage. Still, we may not deduce that a specific type of tube (e.g. the EL-34) has a special frequency response. Again: <u>all</u> tube parameters are frequency-<u>in</u>dependent throughout the audio range. However, cooperating with a special circuit-environment, every tube can and will result in a system that in its entirety is frequency-dependent. The number of circuit variants for power stages in guitar amplifier is not infinite, and therefore findings from the investigation of one amplifier may be applied to some other amplifiers. For example, if a special 6L6-GC sounds trebly in listening experiments with a 4-Ω-Tremolux, it is likely to sound that way in a 4-Ω-Bandmaster because the same output transformer is used in both amps. Even here, though, an imponderability remains in that the impedances of the two loudspeakers may be different, and another one in the fact that even 6L6-GC's sourced from the same manufacturer may be different. Just crowned the test-winner in a Fender-amp, one and the same 6L6-GC may disappoint completely when plugged into a Marshall. Or it may be fine – that depends on the personal taste, the musical style, the specific circuitry, the individual loudspeaker and the individual tube. Blanket-judgments such as "*the KT-66 is a HiFi-tube*" are non-sense, if they seek to refer to resistance against distortion. Because: all tubes were originally developed for HiFi, weren't they?

Fig. 10.5.37 shows the transfer characteristics of typical power tubes. These measurements can (and are supposed to) merely indicate the behavior of such curves in principle - no recommendation regarding which tube to purchase may be derived. For one, because e.g. KT-66's from two manufacturers can differ drastically, and second because even KT-66's from *one and the same* manufacturer can vary in their parameters. In order to obtain a reasonably reliable statistic, numerous tubes would have to be acquired – at more than 50 Euro per piece, this is not a really desirable undertaking.

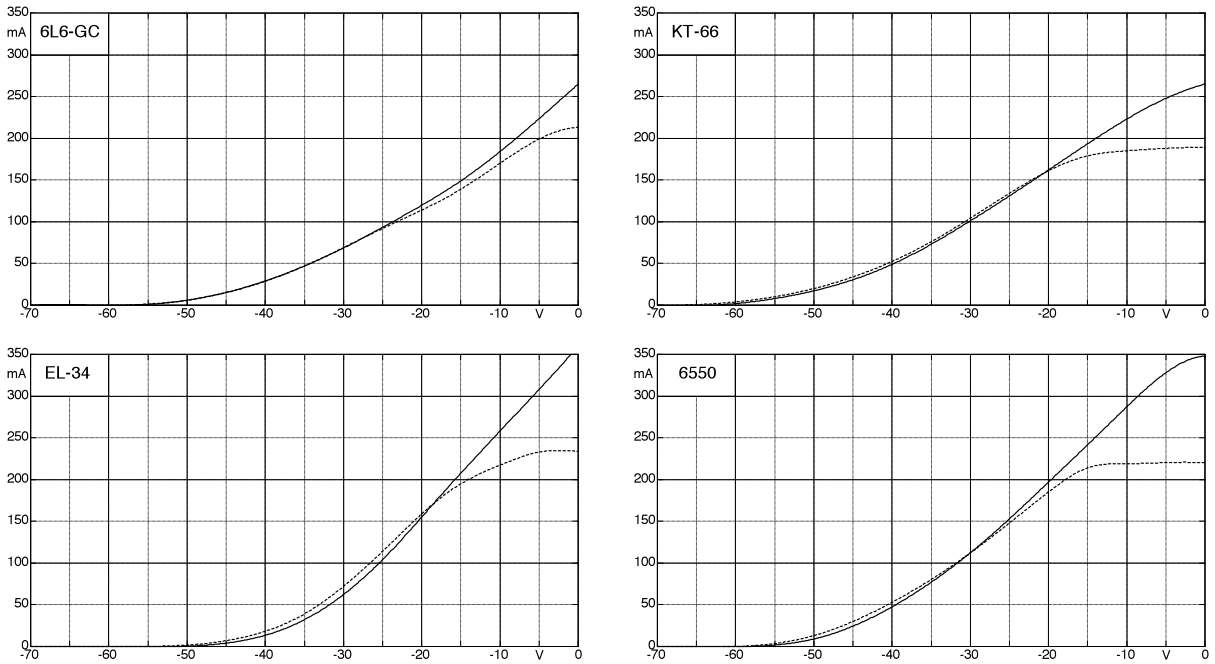

Fig. 10.5.37: Difference in the transfer behavior (measurements), ohmic load 8 Ω (——), 16 Ω (----). M50A.

The measurements were done using a TAD M50A output transformer and an ohmic load-impedance. The differences in the gain (large with the EL34, smaller with the 6L6-GC and the KT-66) are obvious, which is not surprising: the datasheets show the same under the entry "transconductance". We find large maximum currents for the EL34 and the 6550, allowing for inferences regarding loudness, and we see similar curves with the 6L6-GC for both 8Ω and 16Ω, but larger differences between the two load conditions in the 6550. Taking the frequency response of a specific speaker as a basis, we can deduce basic sound-variations from the difference in the respective two curves for each tube. Both curves similar = load-independent, stiff current-source; pronounced differences = less power at a higher-impedance load i.e. less treble. For such statements we certainly need to look into the linear behavior, too; it is only with such an overall consideration that we arrive at a reliable conclusion.

Can these measurements support the notion that an EL34 will distort rather early while the 6550 remains "solid and clean", as Pittman writes in his Tube Amp Book? Before diving into the slightly different 16-Ω-curves, lets first linger and check Pittman's bias-voltages: -50 V for the EL34, and -68 V for the 6550. That does, however, not match Fig. 10.5.37, at all! What could A.P.'s approach have been here? Probably, his reference amplifier works with a higher supply voltage than the one used for the above measurements. An increased voltage at the screen grid could explain a more negative control-grid-voltage, however: how generally valid are Pittman's statements then? His subjective evaluation of the sound shall not be put into question, but his GT-Electronics-Dual-75-Amp is not really that ubiquitous, and at -50 V (EL34) it is not set typically, either. As Pittman notes a few pages later: in a 50-W-Marshall there'd be -43 … -40 V.

Also, another striking aspect in the Tube Amp Book: two tubes (6L6-OS vs. 6L6-B) differ slightly in their distortion: "The 6L6-OS clips a little sooner". Every single tube is, however, also selected according to its distortion characteristic and designated with numbers: 1 – 3 = early distortion, 4 – 7 = normal distortion, 8 – 10 = late distortion. This is because of unavoidable scatter in manufacturing even tubes of *one and the same* type will have different data. Now, does a 6L6-OS with a rating of 10 still distort earlier than a 6L6-B with a rating of 1?? As commendable as Aspen Pittmans's approach towards quantification is, one has to get lost in this vast desert: the circuits are different, and so are the output transformers, the loudspeakers, the tubes – and in any case on top of that the subjective expectations, as well. In a Marshall, Pittman opines, the 6550 sounds *"loud and crunchy"*. If you don't favor that, you change the circuit to install EL34's, because that tube sounds *"smoother, with warmer distortion"*. Turning over a page or two, we also read that the EL-34 may sound *"gritty and a little squashy"*, and the 6550 may sound *"fat and clean"*. Or it may sound *"extremely harsh"*, as we read in publications from German authors. We however also find the latter attesting the EL34 a "*warm sound*" which doesn't really fit "*gritty*". On the other hand, German advertisement has the EL34 giving a "dynamic attack", while the US-colleagues arrive at the verdict: *"The EL-34-setup seemed to lack dynamics"*. For the 6L6-GC, evaluations stretch from "*a fat, more mid-rangy singing distortion sound*" to "*more unstable and mushy*".

In fact, it is not only purposeful but imperative to judge the tone of a sound source according to auditory criteria. Measurements are (hopefully) precise and objective, but they are not necessarily directly linked to our subjectively perception. That's why we carry out auditory experiments. If both too much and too little dynamics are attested to the EL34, several reasons are conceivable. Different amps may have been used, or different music played, or just different EL34's may have found their way into the experiment. Indeed, it seems that everything that sports a glass cylinder of 8 – 9 cm length and 33 mm diameter may call itself EL34. Tube retailers do not angrily send back to the manufacturer all rejects that fail the specifications in the Philips-/Siemens-/Mullard-datasheets, but sell this junk – with a markup – as *"specially selected"* merchandise. Which isn't totally inaccurate, either, somehow. Why didn't others think of that? "*For a premium of an additional 500 Euro you can get a selected TV-set the right-hand screen-half of which remains dark.*" Wouldn't that be a cool idea? It is, for amplifier tubes. It is even legal, because today your EL34 is not just designated as such but it's now called EL-34-SVT, EL-34-Cz, EL-34B-STR, EL-34C, or EL-34R, etc. – and any notice of defects can be averted. That does not mean that none of these tubes meet the specifications; some even exceed it – but some will remain 20% below the given current specification. Others may reach the specified current but fail regarding the transconductance. Apparently, the scatter is big enough for Groove-Tubes to designate one of ten (!) subgroups to each tube. These 10 subgroups will have to be significantly different, too – otherwise e.g. three groups would have been sufficient. Now lets consider, on top of this, that the power tubes are fed by driver-tubes the data of which are also subject to a noticeable spread. Furthermore, the power is supplied from circuits including rectifier tubes that may by called (despite individual "selection") rejects (see Chapter 10.7.4). In view of all this the question "*what does the 6L6 sound like*" can only be answered with a sobering "beats me – no clue – not a hunch". Sorry, folks, thou ask'st the wrong man.

"...I have to point out that my experiments trying to map the sonic differences between various tube-types to sound-files did not meet satisfactory results. The recording/reproduction-process minimizes the differences to a minimum such that almost nothing remains of the described differences. We can hardly conclude anything comparable to what is experienced as a difference when playing." (Gitarre&Bass 6/09).

10.5.11 Match Point

Amplifier tubes run through a multi-stage manufacturing process that does not tolerate any major errors. If a process step does not go according to plan, the tube parameters deviate from the values given in the datasheets. The manufacturers (and some retailers, too) therefore test all manufactured tubes, and weeds out the sub-par specimen. The remaining "good ones" are ennobled with the attribute "selected" and sold to the consumer. However, rumor has it that there were some singular cases in which deficient tubes have found their way to the musician. For this reason, the consumers of the "hot goods" from time to time stage a match in which the "matched tubes" have to compete against each other: comparison tests. These typically are a merry ado, extensively covered in the trade magazines. *Translator's note: From the point of view of the scientist and engineer, these reports often have a rather special "quality" bordering on the dysfunctional. That, however, does not seem to bother the testers nor many of the readers even if the process is repeated in the exact same way – in fact the contrary appears to be the case. Is this testimony to the "magic of the tube"?*

10.5.11.1 Selecting and Matching

Translator's note: I choose not to translate the 1st paragraph here because it is a send-off targeting the often excessive use of English terms in German music trade-magazines. Corresponding German terms would be available, so often English is brought in just to sound cool, or to hide a lack of proper understanding of the subject matter behind impressive English terminology. A translation of this paragraph would therefore almost by definition not work in English. Having said that, isn't a term like "transconductance" just marvelously sexy and seductive if we write about guitar amps? Anyone talking about guitar technology should put "transconductance" to good use in any conversation. Seriously though, "transconductance" is a parameter that tubes are "matched" by – so let's get back to our book …:

Both chemistry and mechanics are involved in the manufacture of tubes, and in both areas, technical tolerances exist. The cathode coating, the metals of the electrodes, the wound grid-wires, the getter, the insulators, the vacuum – varying parameters wherever we look, and therefore all tubes differ in their operational behavior. The really bad ones get to be thrown away, but the parameters of the useable tubes are still subject to scatter. Consequently, they are individually measured ('selected'), and for use in power stages they are paired up ('matched'). It is customary to operate the power tubes in the typical operating point ('at idle') and to specify the plate-current (PC) flowing at a manufacturer-specific supply voltage (e.g. PC = 41 mA). Usually, the supply voltage is not indicated but this is not that necessary if it is typical for the amp (and consistent). It is, however, not sufficient that two tube characteristics coincide in a single point since the tube is subject to a drive signal, and both voltage and current will change accordingly. It is therefore purposeful to check also the dynamic behavior – on top of the static behavior. Enter the transconductance. Barkhausen [Lit.] put his tube-formula together using it: durchgriff (see Chapter 10.5.1) x internal impedance x transconductance = 1. The **transconductance** indicates how strongly the plate-current changes with variations of (only) the grid-voltage. Since the $I_a(U_g)$-correspondence is non-linear, the transconductance can only be determined (as differential quotient) for small drive levels: $S = dI_a \,/\, dU_g$, for constant U_a. The information of e.g. S = 5 mA/V consequently expresses that plate-current changes by 5 mA if the grid-voltage is changed by 1 V. In this scenario, the plate-voltage must not change i.e. the load impedance must be zero – therefore the more extensive, alternate term would be **short-circuit-transconductance**.

Two straight lines are identical if they share one point *and* have the same slope. If characteristics of tubes were straight lines it would be sufficient to measure one point (PC = plate current) and the slope (S = transconductance). However, tube characteristics are not straight lines and therefore two tubes that have been 'matched' via *one* PC- and *one* S-value may very well differ. **Fig. 10.5.38** shows corresponding measurement results. The 35-mA-operating-points of the curves shown in the left-hand section correspond, but the plate-currents of these two tubes differ significantly for grid-voltages converging towards zero. The two tubes documented in the right-hand section show – at the 35-mA-operating point – approximately the same transconductance but their bias-voltage differs. The EL84s on the left are specified with a transconductance of 9 mA/V, the EL84's on the right with 10 mA/V – not a big difference. Given such similar 'matching specs', we would not expect curves differing as strongly.

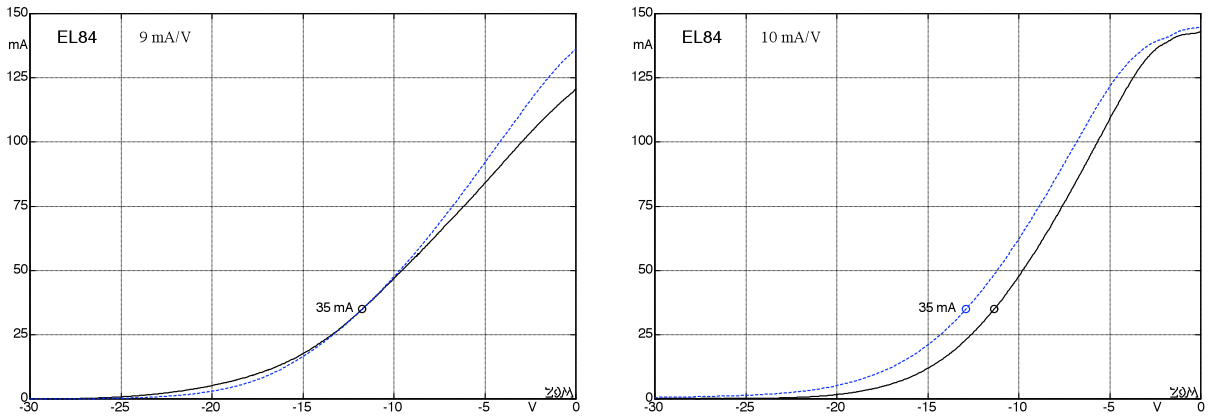

Fig. 10.5.38: EL84-tubes with 'matched' transfer characteristics. $U_B = 350$ V, $U_{g2} = 300$ V, $R_a = 2$ kΩ, $R_{g2} = 0$.

Or maybe we should. Maybe this is why the H&K sales department states: *"for our H&K-amps, we had purchased selected, i.e. matched tubes, but we still experienced a high rejection rate because many tubes did not meet our requirements (Gitarre & Bass 4/09)."* Please note: if you match tubes in only one point of the characteristic, they are not necessarily a match at other points. In practical operation a tube does receive a drive signal, and here not only operating point and transconductance play a role, but among other things also the behavior in the extreme ranges: how well does the tube insulate in reverse operation, how much current will it draw when fully driven, how big is residual voltage caught in the tube. All this should also be tested, shouldn't it? No, not as a rule it isn't – because often there is not even any insight that such measurements would be required. Frequently, the equipment is lacking, too – available is merely your no-name tube-tester indicating "bias" and "transconductance", and that's it: done! To compensate, the plate-current is determined to the tenth of a milliampere, and consequently the PC-values of the 'matched' tubes correspond to the tenth, too. You want to avoid the risk that a musician complains because 36,6/36,7 mA is offered as perfectly matched. Assuming an allowable scatter of the plate-current of ±5 mA, a bin width of 0,1 mA results in 100 different bins. If the transconductance is to be matched with three digits, as well, there might be 100 "transconductance bins". And so the matching person (is he/she a matchmaker, then?) is confronted with 100 x 100 boxes, and bags 'em: every pair matched to a percent. In some cases, this process, tube pairs of astonishing synchronization will go on sale, as shown in the left-hand section of **Fig. 10.5.39**. But then there will also be badly matched ones, like the example given in the right-hand section of the figure. If anyone absolutely is of the opinion that power tubes need to be matched: here you are being served – either way. Incidentally, the two EL34's are, at 43 Euro (per pair), not low-cost but of "excellent quality".

Of course, the transconductance-data printed on the boxes of the measured EL34's correspond to a tenth of a percent: 11.28 mA/V, for both tubes. That may even be correct in some point of the curves – it is, however, unlikely that these tubes were tested at all under full load. And so this "matching" is of little use.

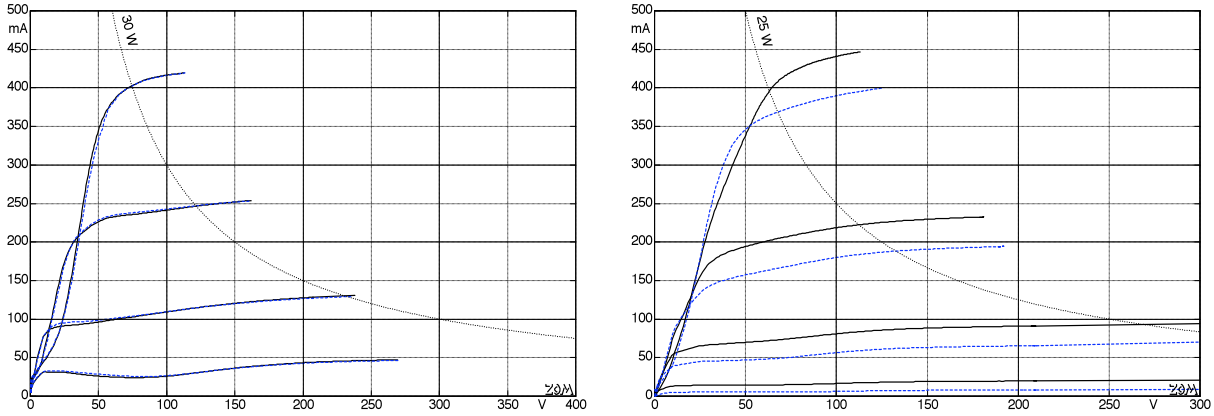

Fig. 10.5.39: Output characteristics of 'matched' tubes. Left: 2x EH-6550, right: 2x EL-34B-STR.

Shelling out e.g. 35 Euro rather than 13 Euro for a pair of EL84's because they are 'selected' and 'matched', it seems only fair to expect not just well matching characteristics but also a correspondence to the curves published in the test certificate. In **Fig. 10.5.40**, we see a comparison between a reference (Philips) and two newly-developed EL-84-STR. According to promo, the latter are supposed to introduce a new standard, and guarantee minimum production scatter. While the idea of a standard may be interpreted this or that way, the fact that the scatter in not minimal in the new tubes is clear enough to be recognized by even the most cloth-eared head-banger. It seems hardly possible for any retailer to more efficiently shoot down his own highly-praised "premium dynamic matching".

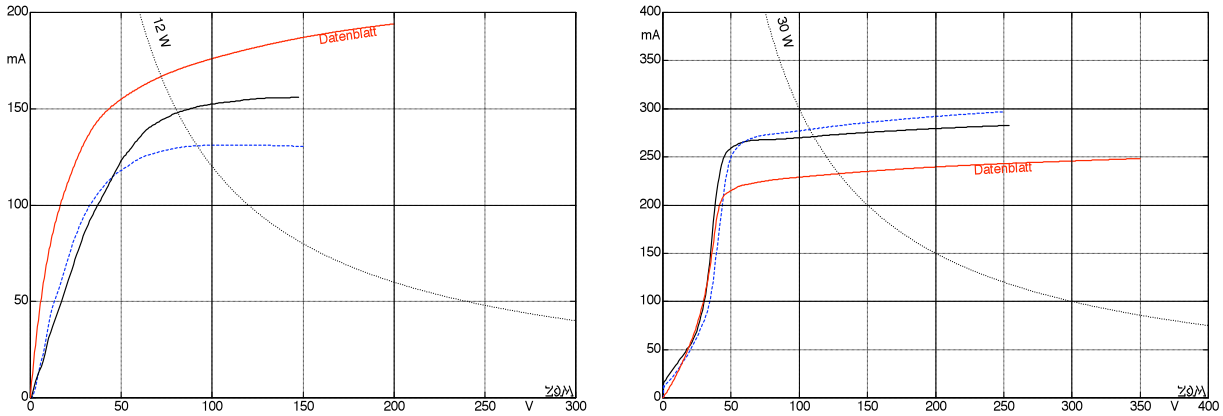

Fig. 10.5.40: Output characteristics ($U_{g1} = 0$) of two EL-84, 'selected' and 'matched'. Right: 2x JJ-6L6-GC.
("Datenblatt" = datasheet)

The right hand part of the figure shows that it is also possible that datasheet-specs are exceeded: both measured JJ-6L6-GC perform better that they need to. That is gratifying, but the screen grids are under more strain, as well, and the gratification about the high performance will quickly vanish if the tubes fail after a short time. That's the gratification of the musician that's gone, cause the dealer will continue to be gratified as the next duet/quartet/sextet/octet/duodectet needs to be acquired. A hint on the side: why not buy, instead of 12 'matched' tubes, 35 industrial tubes for the same money and check whether they won't do the job just as well, either?

10.5.11.2 Comparison tests

Since the market for tubes is entirely non-transparent, every purchase amounts to a gamble. Good to have the "tube gurus" who give recommendations, or even organize **comparison tests**. Which one is the best tube? In many corresponding test reports, the guitar used is presented with much enthusiasm (*1958 faded vomit green*), as is the involved amplifier (*1962 brown Deluxe with Marcus Hotsteam's Mod No. 17*), and the speaker (*a pair of Pinkywinky-Tubbys, with more than 150 h of running-in by playing Blackmore-licks*) so that everybody realizes that true experts are on the job here. Popular is also to point to the jurors (*even Blind Fat Broonzy had a listen*), and the location of the affair (*we all met in Hamburg*) – possibly because of the specific given air pressure. Then there's some dignified unpacking: 2 pieces 6L6-GA, 2 pieces 6L6-WGB-STR, 2 pieces GE-6L6-WGC-NOS (*loaned 'em from Crack Snootshack's pal*), and many another elitist precious'. Plug 'em in, warm 'em up, listen to 'em. "*Most of us arrived at the opinion that the WGB is a touch louder but doesn't give the oomph of the WGC; some liked the GA better, though. Everybody agreed, however, that somehow the NOS very clearly sounded the tightest."* Man, those tube tests – one could get addicted to them. Really informative, somehow.

Not to be misunderstood: this is (so far) a free country; hey, any minister of finance can tell tall tales about his state bank – so why shouldn't aging guitar-slingers just as well publish a tube test or two? Is it sufficient to use a mere two specimen per tube type? Well, at $ 200 per pair, we get that. It is an irrefutable axiom, that listening tests are imperative – just as the fact that never ever will any measurement data be published. As a rule, the tester will have procured that *"Faded Vomit Green"* easily worth 6 numbers among friends – but there is no adequate instrumentation. And even if that were available, the tester could not be bothered to get an understanding of how inter-modulation distortion and difference-tone distortion is not the same thing. Rather, an impulsive "*forget all that theoretical baloney*" will be included in the test-report – and that will not be entirely off the mark, either. Amplifier tubes are designed to be listened to, not to be measured. However, it is the measurement that allows for elegantly objectifying any differences. As a supplement to the listing test, of course. "*Of course NOT*"? Well, it's a free country (see above).

Such listening tests convey the impression that every type of tube has its own sound-shaping characteristic. Indeed, the sound of an amp can noticeably change as the power tubes are swapped – and so each tube must have its special frequency response, musn't it? It will boost of cut the treble, won't it, or it will amplify the bass with particular force. An analogous conclusion would be: as we feed more air to the Bunsen burner, the flame will become hotter – therefore air is combustible. Well, it ain't – and in just the same way, all tube parameters are frequency-independent throughout the audio range. We should not give highest priority to thermal infrasound effects, nor to MHz-effects. As every electronics-undergraduate learns in the circuit-design course: changing a frequency-independent gain-factor in a system with negative feedback may well change the overall transfer function in a frequency-dependent manner. The same can happen if the internal impedance changes by a frequency-independent factor. The frequency response does depend on the tube, but interactively, specific to the amplifier and speaker. Comparison test for tubes are always flawed in that one never knows how far the results are at all applicable to another amplifier. Moreover, one needs to be afraid of a complete and utter disregard of the basic rules of psychometric test-methodology: the test persons are plain prejudiced because no blind testing is done. Or, the test signals are changed in addition to the tube-changes: someone/anyone plays something/anything on the guitar. Reproducibility? No such luck … dream on!

Everyone in the process of purchasing tubes and banking on general verdicts such as "*for distortion sounds, the Sovtek 5881 WXT was the absolute winner in the test*" should know that the parameters of tubes are subject to scatter (due to the manufacturing process). The left-hand section of **Fig. 10.5.41** shows the output characteristics of 6 6L6GC-tubes made by Ultron. That's only 6 tubes and therefore not enough to indicate the maximum scatter that can be expected. This sample is, however, sufficient to recognize that these 6 arbitrarily chosen tubes all manufactured by Ultron vary about as much as the TAD 6L6-WGC differs from the Tungsol 5881 (i.e. two different suppliers!) in the right hand section of the figure.

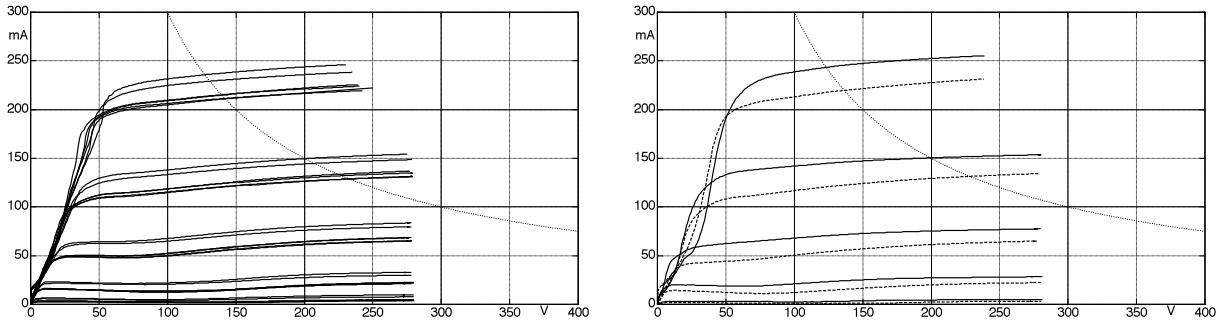

Fig. 10.5.41: Output characteristics: 6x Ultron 6L6-GC (left), Tungsol 5881 and TAD 6L6-WGC (right).

Unfortunately, the "matching" of tubes does not get rid of the problem. A comparison test as mentioned above elucidates: "*moreover, the tube pairs need to be optimally matched. In other words, the idle-current needs to be the same in both tubes as exactly as possible. The pair supplied by Tube Amp Doctor was perfectly in tune. We measured a deviation of only 2 mA. A mismatch of more than 5 mA would cause crossover distortion, and weak and inharmonic sound.*" Figs. 10.5.38 & 39 have already shown that equality in idle-current (bias-current) does in no way guarantee equality in the characteristics. Also, the term "transconductance" does not even show up once in the test report, just as power measurements, frequency responses or characteristic curves are completely foregone. Rather, the insights won are limited to blanket judgments such as "*the KT-66's are, in principle, HiFi-tubes and were deployed in 200-W-Marshall tops.*" Now, that indeed is a surprise. Not so much because 'in principle' all tubes should be HiFi-tubes, but more so because 4 KT-66 can hardly generate 200 W. Michael Doyle writes in his Book on Marshall that KT-88's were used in the 200-W-power-stages – that makes much more sense. And another citation for all students of psychology who need a quick additional example for their exam: "*Steve Ray Vaughan had a quartet of KT-66 in his famous Dumble Steel String Singer amp. Anyone who knows Stevie's album 'The sky is crying' already knows fundamentally how these tubes sound.*" In principle like HiFi, of course, don't they? Once you are aware of that, tube tests become rather dispensable – at least in principle.

There are some indications of this insight filtering through a bit; in a more recent test (Gitarre&Bass 3/2009), we read: "*Another problem in my testing was the possibility of a complete reversal of the results, depending on the amplifier*", and "*occasionally, only little remains of the clear differences that are experienced directly in front of the amp.*" What does remain is least one question: is it possible that a Chinese KT-66 can be "*through and through authentically*" sounding like the old MOV-originals, although its data (at U_a = 50 V) differ by a factor of three (!) from those in the old MOV datasheet? No, this is not proof that datasheets have no connection to the sound: every sound is based on voltages and currents, the correspondence of which is depicted in characteristic curves. And if that weren't necessarily so, we wouldn't have to so carefully match the plate-current, either, now would we?

10.5.12 Special tube power-stages

In the following, a few selected tube power-stages are presented and discussed with respect to some parameters. In doing so, we will always consider that the behavior of every tube amp will depend on its individual components. All measurement curves shown were taken from a specific amp – even if an amplifier of the same type is built according to the same schematic, it can still behave differently.

VOX AC-30

The VOX AC-30 and its predecessor AC-15 are the guitar amplifiers often seen as "the" prototypes for the class-A push-pull power stage. We will not investigate the fact that there was a series of similar amps (e.g. from Gibson) – but we will look into the issue whether the AC-30 actually is powered by a class-A push-pull output stage. Technical literature consistently defines this type of operation via two aspects: the power tubes must not be driven into reverse operation, and the operating point must be located in the middle of the load line. What is the situation lined up in the VOX?

Four EL84 are employed in the AC-30, two each in a parallel configuration to double up the current. **Fig. 10.5.42** shows the output characteristic of this power pentode, with the operating point set to about 310 V / 47 mA – at least for the early variants. After the silicon rectifier had superseded the rectifier tube, voltages of more than 360 V found their way into the amp, but lets pick the "original VOX", the way it was built at the beginning of the 1960's, as object of our investigation. Without any drive signal, we find, at the operating point as given above, a power dissipation of 14 W per tube – mind you, that's 2 W in excess of what the datasheet allows. Still, that is just about tolerable (if we agree to a reduced life expectancy of the tube). However, a symmetric drive-situation (i.e. text-book class-A-operation) is not possible for this operating point: at a control-grid-voltage of about –10 ±6 V, the power tubes start limiting to *one side* of the signal, and therefore the provisional conclusion needs to be: **the VOX AC-30 does not feature a class-A push-pull power stage.**

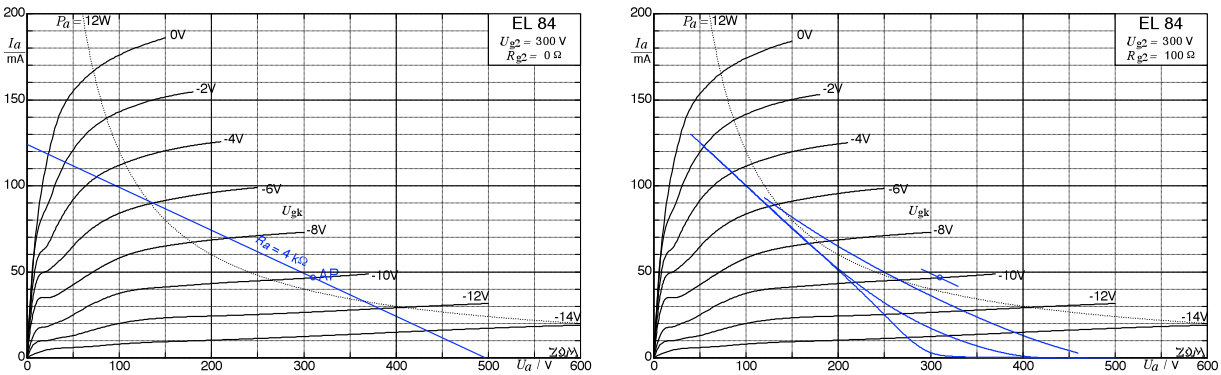

Fig. 10.5.42: Output characteristic of the EL84, ideal load line (4 kΩ) at a supply voltage of 310 V (left).
On the right measurement results for a VOX AC-30 are given (ohmic 8-Ω-load at the 8–Ω–output).

A more exact analysis of the load line confirms this diagnosis (Fig. 10.5.42, right hand section). For small drive levels the expected load line occurs, with a slope resulting from the 4-kΩ-load-impedance. For increasing drive levels, however, the OP wanders off into the lower ranges i.e. to smaller current values, and the slope changes from 4 kΩ to 2 kΩ. This indeed needs to happen, because the setting of the grid-current in the power tubes will polarize the coupling cap (Chapter 10.4.4), and also because each of the tubes now practically works in push-pull class-B mode (Chapter 10.5.3 & 10.5.5). If the AC-30 power stage indeed

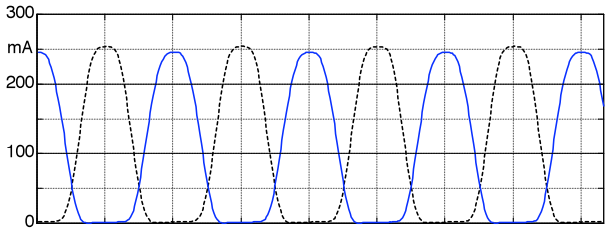

Fig. 10.5.43: Primary currents in the output transformer.

were a class-A push-pull circuit, current would have to flow during the *full* signal period. That is not the case, however, as the measurements shown in Fig 5.4.43. clearly prove. High drive-levels let current flow in the power tubes only during *half* the period, and therefore the AC-30 power stage is not a class-A power amp.

It is astonishing how tenaciously the fairytale about the allegedly unique push-pull class-A power stage keeps being repeated. In the book about VOX [Petersen/Denney 1995], this starts already in the introduction written by Brian May: *"The VOX AC-30 ... uses a Class A configuration."* Co-author Denney should know better: he has developed this amplifier, after all. The same with tube-vendor TAD: "*The sound of the class-A operation has been made legendary by the VOX AC-30! Class-A operation yields several advantages: a thick, 'three-dimensional' sound with pleasant, slight compression, singing sustain and harmonic, controllable distortion are typical*". Aspen Pittman opines in his collection of schematics: *"Contributing to the amp's smooth tone in both the clean and distorted modes is its very unusual Class A circuit designed by Dick Denney"*. Well, this power-stage circuit was not that unusual: two power pentodes, a common cathode resistor (i.e. automatic generation of bias-voltage) – we can easily find that years before in Fender amps (e.g. the Deluxe 5B3), and in Gibson amps (e.g. the GA-40); this was textbook-standard. It was only the value of the cathode resistor that varied – it set the **operating point** and made for a "hotter" or "cooler" operation of the amp (Chapter 10.5.8). And indeed, here the VOX does show a peculiarity: it operates at the hottest possible tail-end, with a power-dissipation of 14 W (average value) at the plate (the datasheet gives a maximum value of 12 W). However, a hot operation (or cathode resistor, respectively) does not automatically imply push-pull class-A.

Why push-pull class-A in the first place? To get the least non-linear distortion! Due to the superposition of differently-curved tube-characteristics, the non-linear components compensate each other, the THD decreases. Literature explicitly points out, however, that this only holds for triodes: *"For pentodes the push-pull A-circuit does not yield a significant improvement relative to the THD of the single tube [Schröder]"*. As a reminder: the EL84 is a pentode! The literature has more to in store: *"In a correctly balanced push-pull A-amplifier a capacitor is <u>not</u> required to bridge the cathode resistor. In a AB-amplifier it is, though [Langford-Smith]"*. The AC-30 does possess such a capacitor. Lastly: we find the voltage at the cathode resistor specified in the VOC-schematic; it is 10 V without input signal, and 12.5 V at full drive level. If this were a push-pull-A-circuit, this voltage would remain constant. An old **AC-15**-schematic from back in 1955 reveals a common cathode resistor amounting to 130 Ω, and R_{aa} = 8 kΩ for the load impedance at the plate. The Siemens-datasheet (from 1955) recommends, for a plate-voltage of 300 V, a common cathode resistor of 130 Ω, as well as R_{aa} = 8 kΩ. Coincidence? Of course not – the circuit designers were wise enough to follow the recommendations of the tube manufacturers. Siemens, Telefunken, Philips – they all specified R_k = 130 Ω and R_{aa} = 8 kΩ for the EL84-push-pull power stage. No, not for push-pull class-A configuration! These recommendations from Siemens, Telefunken and Philips are given for **push-pull class-AB configuration.** The AC-30 included four EL84 instead of two, i.e. double the current, and thus half the value of the cathode resistor. Old plans show an R_k of 80 Ω to begin with, but it soon was reduced to 47 Ω. Half of 130 Ω would have been 65 Ω – and so they opted for slightly higher output power (and slightly less tube endurance).

Also, the decision had been taken to do without any negative feedback (NFB). The typical Fender amplifier from the late 1950's fed back a part of the output signal to the input of the phase-inverter and reduced the non-linear distortion of the power stage that way. The AC-30 (from 1958) dispenses with that kind of negative feedback; for this reason, some presume that the distortion in the AC-30 would be "extremely high". Well, it's not – as **Fig. 10.5.44** shows. Granted, 5% THD is not exactly studio-standard, but the AC-30 was never intended to power studio monitors. At small drive levels, the harmonic distortion is a low as k_3 = 0.3%, and with increasing drive levels, the distortion gradually rises. This is in contrast to power-stages that feature strong NFB and, correspondingly, a sudden increase of the distortion at the drive-limit. Apparently, VOX-guitarists prefer the gradually rising distortion.

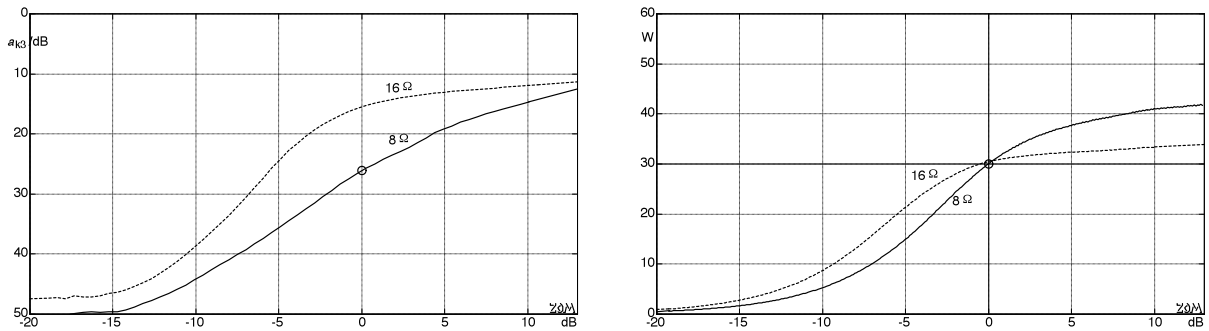

Fig. 10.5.44: AC-30, 8-Ω-output: distortion below signal (left), power (right); Abscissa referenced to P = 30 W.

It has already been mentioned that tube power stages cannot be described with *one single* characteristic curve because the operating points shift due to re-charging effects. **Fig. 10.5.45** shows corresponding measurements taken with varying drive levels. With increasing drive, the transmission characteristic flattens out, with a saddle-point appearing in the origin. For a load of 16 Ω (right hand part of the figure), the curves generally run steeper (high internal impedance ≈ current source). The flattening of the curves can be interpreted as a kind of compressor that reduces the gain of the power stage as the signal level increases. The load-dependency of the output voltage results in emphasizing the loudspeaker resonance and the high-frequency signal-components (compare to Chapter 11). In contrast, a power stage with strong NFB would have a drive-level-independent, sharply bent characteristic similar to the one discussed in Chapter 10.1.4. The maximum power-yield merits some attention, as well: with a stiff voltage-source (low internal impedance), the power-limit for a 16-Ω-load would be half of that for an 8-Ω-load ($P = U^2 / R$); the AC-30, however, reaches more than 80%.

Fig. 10.5.45: Characteristic curves of an AC-30 power stage. Left 8 Ω, right 16 Ω load (at the 8-Ω-output).
These figures are reserved for the print-version of this book.

This special power-stage characteristic is also documented by sweep measurements. For the latter, the AC-30 was connected to a Marshall 1960-AX – not your typical AC-30 speaker but able to take significantly more punishment than the fragile and overly expensive blue Celestions in the combo. **Fig. 10.5.46** shows the voltage level measured at the 16–Ω–output, on the bottom for small drive level and on top for overload.

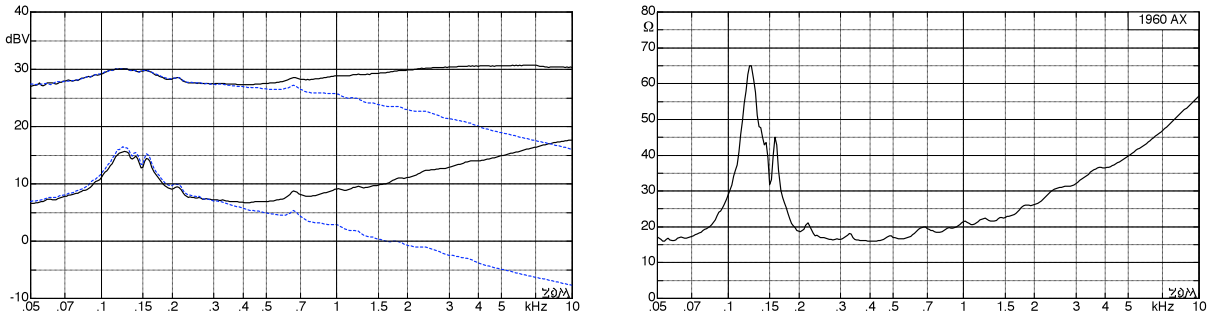

Fig. 10.5.46: Frequency response of an AC-30 power stage, 16-Ω-output loaded with 1960-AX,
Cut CCW (---) , Cut CW (——). Right: loudspeaker impedance (in a reflecting room). Compare to Chapter. 11.8.

It has been repeatedly noted that the power tubes deployed in the VOX do not only suffer when the amp is overdriven, but are under strain already with no input signal present at all. In **Fig. 10.5.47**, we see the power dissipation at the plate and at the control grid for drive levels rising by 30 dB. Without input signal, the power dissipation at the plate is about 14 W in each EL84. The strain on the plate decreases as the drive level rises, and after switching off the input signal there is a short peak in the strain. At idle, the **strain at the screen grid** is just below the allowable limit; with an input signal present, the limit value is very easily exceeded, especially with a high-impedance load (for typical loudspeaker impedances see Chapter 11).

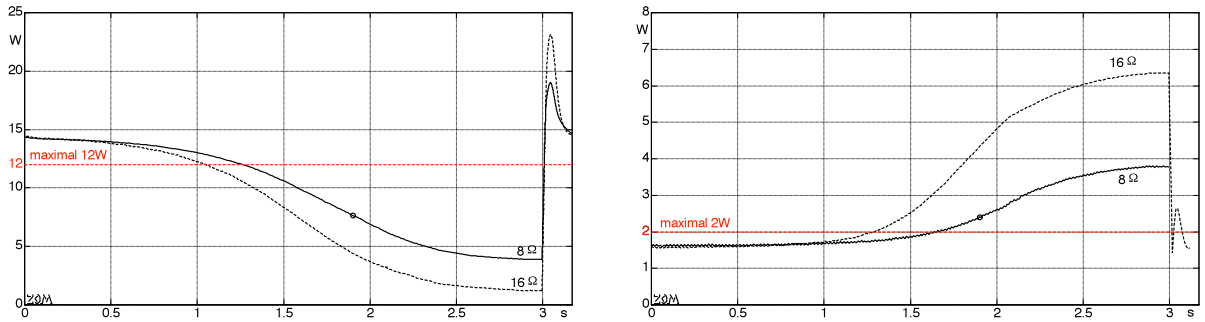

Fig. 10.5.47: AC-30: plate dissipation (left), screen-grid dissipation (right). The level of the sine tone at the input linearly rises by 30 dB from 0 … 3 s, at t = 1.9 s nominal power is reached for an 8–Ω–load. At t = 3 s the input signal is switched off, subsequently there are balancing processes in the capacitors of the power-stage.

The measurements for Fig. 10.5.47 were taken with an AC-30 that had a tube rectifier (GZ-34) in its power supply. Replacing the GZ-34 by silicon diodes will lead to an increase of the strain at the plate in idle to about 17 W; the peak after the switching-off reaches 30 W. The maximum stain on the screen grid exceeds 6 W for an 8-Ω-load, and 10 W for a 16-Ω-load! Since real loudspeaker impedances (including the so-called 16-Ω-speaker) can become higher than 16 Ω (Chapter 11.2), even stronger overload needs to be expected.

The **power supply** merits attention for another reason: the operating voltage for the power pentodes is directly taken from the cathode of the rectifier tube – the voltage has a corresponding ripple. This is not a problem at small drive levels, but it is for strong drive levels since clearly noticeable amplitude modulations result.

Abb. 10.5.48 depicts the time-function of the output voltage, with the 8-Ω-output loaded with a purely ohmic 8-Ω-load. As long as the output voltage is not limited, any fluctuations in the supply voltage represent a (superimposed) common-mode interference that is largely suppressed by the output transformer – the output voltage of the transformer remains un-modulated (left hand section of the figure). In overdrive-operation, however, an asymmetric limiting appears: the maximum plate-current depends on the supply voltage while the minimum plate-current does not (it is practically zero). As a result, an envelope depending on double the mains-frequency is generated – it may be seen, as a first approximation, as a 100-Hz-amplitude-modulation (AM). It is not that a 100-Hz-tone is superimposed onto the input signal; rather, the latter is changed (modulated) in its amplitude. The **envelope** over time (which does not actually exist but is an imaginary auxiliary line) is shown dashed in the right-hand section of the figure.

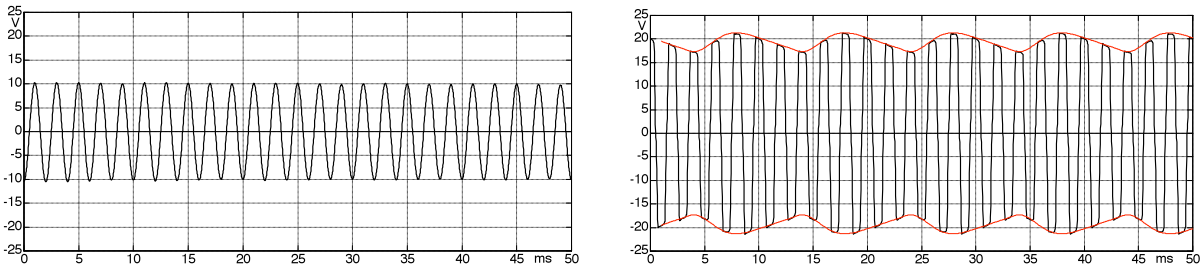

Fig. 10.5.48: Voltage at the 8-Ω-output for an 8-Ω-load. Drive level: half (left), overdriven (right).

Spectrally seen, this 100-Hz-modulation does not make itself felt as a line at 100 Hz. Rather, modulation-lines *next to* the signal-lines result. As a model, the AM can be illustrated as the multiplication 'signal x envelope', corresponding to a convolution in the frequency domain. Since the envelope is not of an exact sine-shape (Chapter 10.7), we not only get a single pair of additional lines (±100 Hz), but several pairs. The level-spectra related to Fig. 10.5.48 are shown in **Fig.10.5.49**: in the left-hand picture, the modulation lines (lateral lines) have a level of 45 dB below the carrier – the (3rd-order) distortion (at 1.5 kHz) is 1%. For the overdrive operation chosen in the right-hand section of the picture, the 3rd-order distortion amounts to 20% with the level-distance between the modulation lines having decreased to 27 dB.

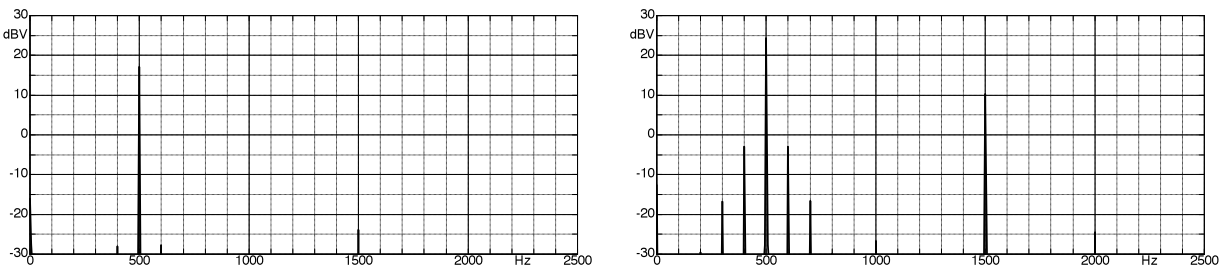

Abb. 10.5.49: Level-spectra related to Fig. 10.5.48.

Not every amplitude modulation that can be measured is necessarily audible – the AM shown in the right hand section of the figure can, however, be assumed to be noticeable as an additional roughness. In the area of psychoacoustics [12], the term "**roughness**" designates auditory perceptions created by fast signal fluctuations that could be labeled as a kind of buzzing sound. Measurements of harmonic distortion normally do not encompass modulation distortions; therefore, dedicated measurements are necessary for this type of distortion.

Marshall JTM-45

James Marshall opened his drum shop in London in 1960, and soon began to sell amplifiers alongside the drum kits. First, the rather expensive Fender amps and others, but from 1962 also the first Marshall amps that his technician Ken Bran assembled as close copies of the Fender (tweed) Bassman from 1959. Young guitarist James Marshall Hendrix was a customer, and both had laid the foundation of their respective careers: one went by the name Jim Marshall, the other called himself Jimi Hendrix. It was Eric Clapton, however, who first attracted worldwide attention to the Marshall amp. He recorded, together with John Mayall, an album the cover of which included a picture showing in the background a Marshall combo: the legendary JTM-45, with model number 1962.

What's so special about the JTM-45 power stage – what is it that creates the legendary sound? A sound – as Gitarre&Bass 7/06 puts it (for once not onomatopoetically but tribo-poetically[⊕]) – of a *"fat and creamy crunch-tone"*, but *"never a Marshall-typical distortion sound"*. Excuse me?!?! A non-Marshall-sounding Marshall? Although we are told that the JTM-45 includes *"all the ingredients responsible for the plexi-sound[♣] that achieved legendary status later"*? Presumably, you can see these ingredients – but you can't actually hear them. 18 months before (G&B 2/05), the JTM-45 was described as *"even hotter and more aggressive"*, and 6 months before (G&B 2/06) with *"clear and fat, with a soft spectrum in the mids"*.

Clearly, fat may be hot – why not. How this hot-fat sound originates is subject of innumerably speculations. It starts with **Clapton's Les Paul** for which pertinent literature holds in stock a vintage of '58, '59. or '60. Shouldn't that be all the same to us? No way – that makes for one heck of a difference: after all, the frets changed over these production-years (they got wider), the neck angle also (it increased), and the cross-section of the neck, as well (it got more narrow). All this should be, of course, "tone-affecting", shouldn't it? And so we would expect E.C. to answer the question which model he bought back in the day (in June 1965 according to G&B 9/08) with an immediate: the '58, of course, because of the big neck that – as we all know – improves richness of tone and sustain [G&B Gibson-Special]. However, he does not answer anything of the like but merely notes: *"No idea"*. No idea? Geez, Eric (as musician, one is on a first name-basis right away), you should know that: the increased neck-angle of the '60-Paula (as this type of guitar is designated in circles of experts) alone would have ruined sustain, and the thin neck of the '60 "*has no acceptable vibration characteristic whatsoever [G&B 3/97]*". Very strange that Eric does not remember. Thank Eric we do have recordings surviving from those "Clapton is God"-times – Beano and such – so we should easily be able to pick out what the deal was. Here's the latest level of knowledge: *"Today the general opinion is that the guitar concerned was a '60-model since both Clapton and Peter Green describe the 'slinky' neck [G&B 9/08]."* Indeed these are tough times for guitar experts: on the one hand they continuously are required to explain that the smallest details of a Les Paul (varnish, frets, neck, pots or tone-caps) have an immense influence on the sound, but on the other hand there is not anyone in the world who could recognize, on the basis of these sound-specifics, and from listening to the Bluesbreaker-LP, the version of the guitar. The stopgap solution then is to reason the guitar-type from on memories regarding the neck-profile. What an odd, make-believe world of Gods and experts … and stopgaps.

Well then: we don't know any specifics about the guitar, but the amp is known: a **JTM-45, 2x12-Combo Type II**, in all likelihood fitted with alnico-speakers (G&B 9/2008).

[⊕] Tribology = teachings of friction and lubricants

[♣] Hopefully, the plexi *will* sound like a Marshall – or still not, either??

In all likelihood? Again, nobody knows exactly. Alnicos will yield *"particularly sweet and harmonically rich treble"*, that is known the latest since G&B 8/05, and so we should be able to hear from the Beano-LP whether alnicos or ceramics are at work. But again, the LP denies any analysis, and although ceramic-powered speakers sound *"entirely different"* compared to alnicos, nobody can pick out from the record which speakers were recorded. Unfortunately, it was just in those days that Marshall started to switch over to the ceramic-Celestions so that both types would be eligible. Again, let's ask Mr. Clapton – he should … no? Again, no memory? Well-well, dear Eric: did various substances abound that much already back then? Indeed?! Then we shall not insist. And on to a look into the expert literature: *"Because Clapton ran the amplifier at full volume, the Alnicos may have been damaged. He may have replaced them with the higher wattage, ceramic magnet Celestion Greenbacks."* This is the voice of Premier Guitar (February 2008). Clapton replacing his alnicos by ceramics? His alnicos, those that will produce – according to Premier Guitar – *"sweet warm tones and a smooth midrange"*? And that generate, according to G&B, *"particularly sweet and harmonically rich treble"* and do *"sound harmonic and with a bite"*. Entirely differently then, compared to the subsequent ceramic-Celestions that yield *"plenty of midrange crunch"* but *"...sounded very different from the Alnico type speakers used in other Marshalls [David Szabados]."* Of course nobody knows whether Eric did actually change the speakers: *"He may have"*, and he himself can't remember. That should be not a real problem, though, because there is that LP, and from it we should be able to pick out the speakers due to: *"sounded very different."* It remains difficult, though, because on the one hand the ceramics sound somehow very different – but not really, on the other hand, because otherwise we would be able to pick them out. In conclusion: we don't know anything in detail about the speakers, either.

We do know one thing, though: the **output transformer** was sourced either from Radio-Spares, or from Drake. That is certain: either / or. It is also known with certainty that the two transformers were not equivalent: the Drakes were *"rougher and more distortion-happy, more mid-rangy, darker than the R.S."*. Unfortunately we cannot pick out from the recording which transformer was on duty for Mr. Clapton, and therefore the retrofit-supplier offers replicas of both transformer, just to be safe. They cost about 250.-- USD (that's for one, not for both), thank you very much, plus customs and shipping, and there you are, another step nearer my God to Thee. You gotta understand why these transformer are so expensive: hand-made! Encouragingly, the core-sheets are not sawed out with a jigsaw – that would have made them seem a bit overpriced. Around 300 USD, that's o.k. – it's a detailed copy of the Clapton-gear, after all. In all likelihood – because we still do not know whether Drake or RS, and moreover the resident expert at G&B offers yet another variant: Mr. Clapton may have operated a pair of speakers having (in conjunction) an impedance of 8 Ω from the 16-Ω-ouput of the transformer. That's a factor of two, so a 100-%-mismatch – or is it 50%? Sorry, it is not easy to theoretically get a handle on these things, so we better draw some conclusions: Clapton's JTM-45-sound is legendary, we all agree on that. If you want to copy that sound, you acquire either a '58- or a '59- or a '60-Les Paul (allegedly differing audibly in sound), fit your JTM-45 with either a Drake- or an RS-transformer (allegedly differing audibly in sound), and install two alnico- or two ceramic-Celestions (allegedly differing audibly in sound) – and now you should firmly, certifiably reside in the midst of Beano-tone. Wow!

Clapton's Bluesbreaker-sound is great – how it originated is uncertain. Readily overlooked: a guitar player was involved of for the time extraordinary skill and talent, and of course studio technology will have had an influence. Clapton's Marshall-combo has disappeared – its specs are unknown. What remains is to use schematics and replicas, knowing that a schematic does not document all details. In the following we will analyze what the hand-drawn sketch in Doyle's Marshall-book reveals.

Marshall's (or rather Bran's) first amplifier was the JTM-45, with two **KT-66's** operating in a push-pull class-AB configuration in its power stage (with a few exceptions). For this mode of operations, the GEC-datasheet specifies an output power of 30 W. The number "45" after the JTM therefore is not an indication of the RMS-power but just promises a seemingly 50-%-advantage over the AC-30. The JTM-45 power stage includes a negative feedback which is relatively strong for a tube amp, with several consequences: non-linear distortion is reduced, loudspeaker resonances have less of an effect, and the amp may oscillate in the RF-range, in particular with the presence control turned down. Feedback functions as **negative feedback (NFB)** if the signal led back to the input is added to the control signal in opposite phase. In the high-frequency ranges, however, phase-shifts may occur (e.g. in the output transformer), and the negative feedback can turn into positive feedback: the amp will oscillate. These oscillations may only happen in a certain range of the drive-signal range where the specific gain and phase-shifts (both being drive-level-dependent) make for a loop gain of larger than 1. It is necessary to avoid such oscillations even if they are located in an inaudible frequency range to begin with: first, because they result in the operation of an illegal RF-transmitter, and second, because they put unnecessary strain on the power stage.

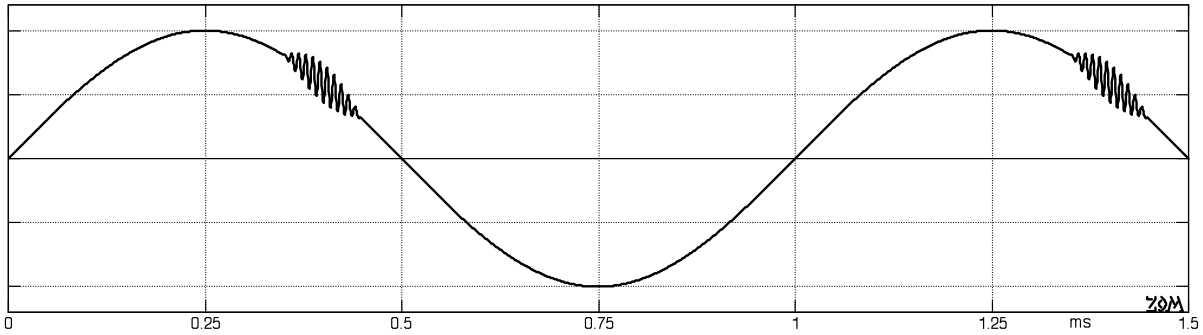

Fig. 10.5.50: 1-kHz-tone with superimposed RF-oscillation.

Fig. 10.5.50 depicts, in principle, the shape of an "RF-infested" audio signal. The RF (often around 150 kHz) is not always recognizable as a clean oscillation – it may result merely in a widening or a smearing of the curve of the audio signal. Small capacitors may be found in the circuit as a "brute-force bug-fix", soldered-in at "appropriate locations" to squash the malady. Much better would be a textbook RC-compensation reducing the loop-gain at high frequencies without adding significant phase-shifts (a LP with limited, defined attenuation at high frequencies). Sure, this is not a trivial topic because with every tube-replacement, the condition for the oscillation is newly negotiated – on top of also being dependent on the loudspeaker. Those who want to address this issue in a somewhat less sophisticated manner find a cooperative partner in the form of the Presence control. Just turn it to the right (up, CW) and the annoying RF is gone. It may be surprising that increasing the gain at high frequencies will choke the oscillation – however this reduces the loop gain that determines the tendency to self-oscillation.

Besides the signal-feedback via the NFB-network designed into the amp, another factor may support RF-oscillations: capacitive coupling of non-shielded components. In fact none of the components found internally in a Marshall are shielded, which is why even the position of individual wires can co-determine the tendency to oscillate. Incidentally, this is another detail that cannot be found in the schematic.

With the JTM-45 power stage operating in class-AB mode, there is, besides the choice of tubes, another degrees of freedom: the bias-current (or the offset-voltage at the grid). It would take us too far afield to show all significant characteristics for all appropriate tubes at several bias-current-settings, and therefore just a few examples shall do. **Fig. 10.5.51** shows measurements of the harmonic distortion (a_{k3}), without NFB in the power stage (left) and with NFB as found in the original circuit. In his comments regarding the Marshall circuit, Ken Bran does not make any secret of the fact that the Fender 5F6A-Bassman was used as a model. Therefore, it is not surprising that in both amplifiers, a 27-kΩ-resistor feeds back the signal to a 5-kΩ-presence-control. However, in the Bassman the 2-Ω-winding of the output transformer is the source, while in the JTM-45, the 16-Ω-winding is tapped for this. The negative feedback in the Marshall therefore is three times as efficient (impedances are transformed with the *square* of ratio of the windings). Whether this was by chance, or due to ignorance, or intentional … who would know 50 years later? In any case, for the successor-models of the JTM-45, the degree of NFB was reduced again – for whatever reason.

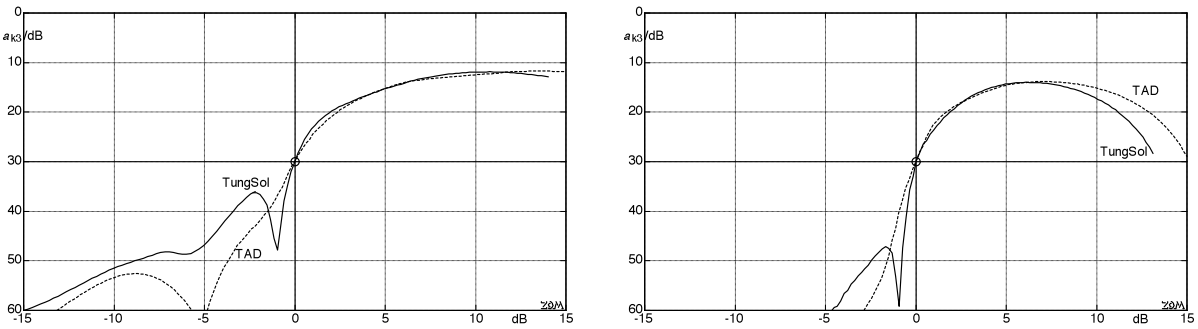

Fig. 10.5.51: JTM-45, harmonic distortion without (left) and with (right) negative feedback in the power stage. An 8-Ω-resistor was connected to the 8-Ω-output for the measurements, R_{aa} = 8 kΩ, f = 500 Hz.

In **Abb. 10.5.51**, the abscissa is set such that at 0 dB and for the specified loading, the signal is just starting undergo limiting. A THD < 1% (i.e. with the generated harmonics 40 dB below the signal) are surely irrelevant for the auditory perception – presumably, 30 dB difference (i.e. 3% THD) would still be inaudible in a guitar amp. There is no binding limit value, though, because too many parameters decide about the audibility of nonlinear distortions. At first glance, the JTM-45 power stage distorts similarly to a transistor power stage – due to the strong NFB. It remains practically distortion-free[⊗] for the non-limited signal, and shows textbook increase of harmonic distortion above the drive-limit. This sentence should in fact be carved in stone: "Marshall's power amp distorts like a transistor power stage" – considering that all those amplifier gurus keep praising the specially-bred Marshall distortion! But then, where should we find something special when we have a copy of an American amp the circuit of which was taken from tube-manuals? The JTM-45 power stage includes a textbook differential amplifier as phase-inverter, two beam-tetrodes with a textbook drive-scenario, and an output transformer as it was offered to a clientele that we would call "hobbyists".[♣] It must not surprise us that secret forces are entrusted to these "Radiospares-Deluxe-Transformer" by its fan base – it is, after all, in the sacred company of Ken Bran's special solder the atoms of which always automatically redirect themselves towards Hanwell. Caution, though, dear buddies: after lugging the amp around you gotta wait for 4 minutes – as we learn in the chemistry course, tin and lead have 4 valence-electrons … they are the so-called inert (passive) heavy metals, and the redirecting of the atoms will take a little while.

[⊗] 60 dB level difference between the generated distortion and the signal corresponds to a THD 0,1%

[♣] In fact, many guitar amp designers had/have a background in ham-radio.

Are you startled yet? Relax, there are, after all, differences to a transistor amp – we must not conclude a general equivalence from the similarity of two distortion-curves. The combination of power tubes and output transformer results in a special power- and distortion-characteristic that in this manner cannot be found in transistor amps. In the JTM-45, the two KT-66 operate towards an output transformer with a rather high primary impedance. Presumably, it first was the RS-Deluxe transformer – nobody can remember which transformer resided in Clapton's "Bluesbreaker"-amp. A: it's been a long time, and B: RS was not a manufacturer but a retailer, and therefore several manufacturers are possible. An American retrofitter (who is not adverse to selling his replacement-transformers) surmises that, in the original JTM-45, an RS-Transformer with R_{aa} = **6.6 kΩ** was at work. The magazine Gitarre&Bass supposes an RS-Transformer with **8.0 kΩ** included in the amp (7/2006), but also considers an 8-kΩ-Drake to be a possible candidate (9/2008). Why would there be such high impedances? The RS-transformer used to begin with was an all-round device intended for applications as universal as possible. Consequently it offered four different primary connections: for KT-66 and EL34 with additional ultra-linear connections R_{aa} = 6.6 kΩ; for 6L6, 6V6 and EL-84 R_{aa} = 8.0 kΩ, or R_{aa} = 9.0 kΩ. The Marshall JTM-45 did not have the ultra-linear configuration, but the KT-66 with R_{aa} = 6.6 kΩ or 8.0 kΩ is today seen as historically correct. By the way, what does the KT-66 datasheet specify? We find R_{aa} = 7 kΩ (ultra-linear), or 8 kΩ for the regular class-AB power stage; for both versions with a cathode-resistor, though. The JTM-45 did not include such a resistor! For this bias-variant, the KT-66 datasheet specifies **5 kΩ** but the supply voltages do not entirely match. Conclusion: neither the output-transformer manufacturer nor the tube manufacturer supplied any exactly matching guidelines to the Marshall developers. Anything else is speculation.

Measurements of the family of output characteristics show that R_{aa} = 8 kΩ is not really conducive for an instrument amplifier (**Fig. 10.5.52**). With a load of 8 Ω connected to the 8-Ω-output, the load line meets the output characteristic of the KT-66 at rather too low a point. In our example, the KT-66 has a scarily high residual voltage but that is another matter. With half the load impedance (right-hand section of the figure), we would close in much better on the ideal condition – and so the conclusion is: for the KT-66 in the JTM-45, ***R_{aa}* = 4 kΩ would be optimal**. That is, at least if a high-power yield is requested. For minimal harmonic distortion, higher primary impedances could be considered, too … but in a Marshall? A 4-Ω-load at the 8-Ω-output would be approximately equivalent to an 8-Ω-load at the 16-Ω-output, a variant also thought possible in Clapton's amp by G&B (09/2008).

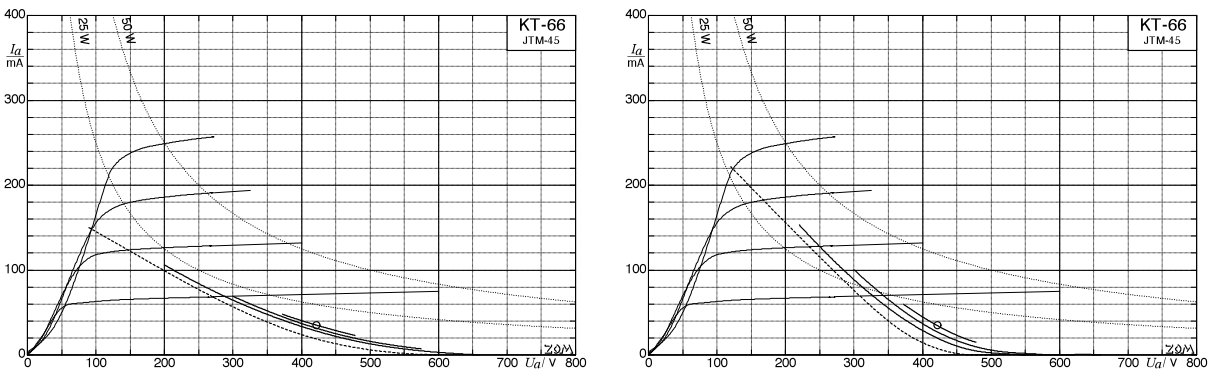

Fig. 10.5.52: Load characteristics, R_{aa} = 8.0 kΩ, 8-Ω-output at an ohmic load of 8 Ω (left), and at 4 Ω (right).

In its measured data, the TungSol-KT-66 corresponds approximated to the GEC-datasheet, while the TAD-KT-66 fails to deliver the required power due to its excessive residual voltage. On the other hand, the latter distorts somewhat less as already shown in Fig. 10.5.52. It is, however, not possible to say how long these evaluations hold: such data change too often.

Fig. 10.5.53 shows the output power of the JTM-45, dependent on the level of the input signal. Fitted with the TAD, it struggles to climb over the 30-W-mark even when overdriven (and loaded with the nominal impedance), while with the TungSol-tubes, it is o.k. to call it a "30-W-amp". Only when overdriven, and with a mismatched load, it gets close to 45 W. Its daddy, the Fender Bassman, could to offer more (Fig. 10.5.62), until Marshall later outperforms it again using the EL34.

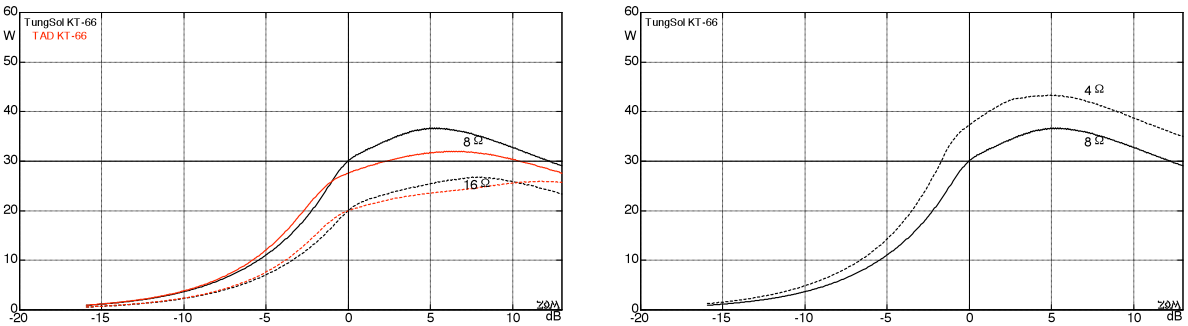

Fig. 10.5.53: JTM-45, output power at the 8Ω-output, loaded with 16 Ω, 8 Ω and 4 Ω, purely ohmic.

In **Fig. 10.5.54** we see the frequency response of the power-stage loaded with a real loudspeaker. With the presence control turned down, the characteristic is almost frequency-independent, despite the frequency-dependent load. As the power stage is overdriven (30-dB-curves) the presence control looses its effect. While the JTM-45 has strong negative feedback (NFB), this apparently was not seen as "the" secret of the Marshall-sound – otherwise it would have been retained in later models. But just that does not happen: rather, the NFB-tap drifts from the 16-Ω-winding to the 8-Ω-winding, and later even on to the 4-Ω-winding; at the same time Marshall increases the feedback resistor from 27 kΩ to 47 kΩ and later even to 100 kΩ. Both these changes reduce the NFB – that, however, affects the successors fitted with EL34's.

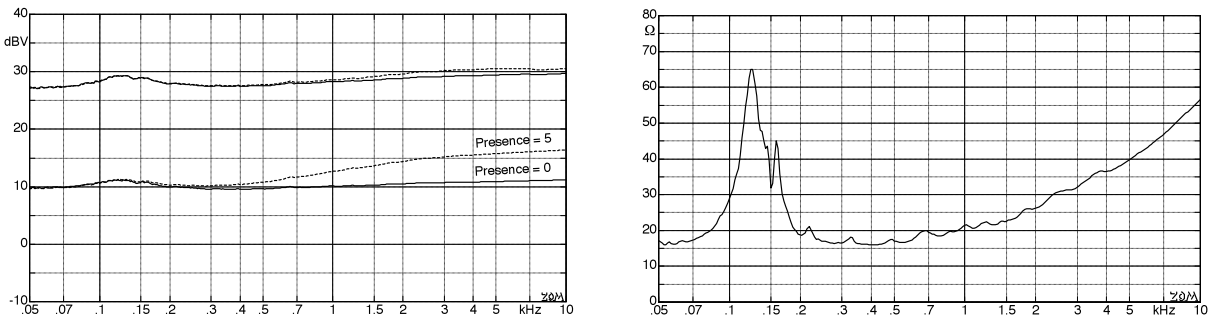

Fig. 10.5.54: Frequency response of a JTM-45 power stage, 16-Ω-output loaded with a 1960-AX speaker (left). Magnitude of the loudspeaker impedance: 1960-AX measured in a room with reflecting surfaces (right).)

We had seen in Fig. 10.5.52 that an 8-kΩ-transfromer does not really challenge the power tubes much. For the power-curves shown in the following, we therefore used a **4-kΩ-transformer**. Connecting the latter, at its 8-Ω-output, to a 16-Ω-load, we arrive approximately at the original conditions (R_{aa} = 8 kΩ). With an 8-Ω-load, the load line just about meets the "knee" of the output characteristic of the tubes (R_{aa} = 4 kΩ). Chapter 10.5.9 already illustrated the effects of such load changes: the smaller the load impedance, the higher the strain on the plate; the larger the load-impedance, the larger the strain on the screen grid. Valid for the JTM-45: R_{aa} = 8 kΩ is the presumed original value; R_{aa} = 4 kΩ would be optimized in terms of the power yield.

The strain on the power tubes is shown in **Fig. 10.5.55** for an ohmic 8-Ω-load at the 8-Ω-output. As the drive level mounts, the strain on both plates first decreases. Then, however, the strain on one of the two power tubes rises again. At the moment the input signal is switched off, we see a high peak in the strain resulting from charge-balancing processes in the coupling capacitors. Since this peak only has a short duration, it is not particularly dangerous to the tubes. Conversely, the screen grid is in more danger: as soon as the power stage is overdriven, the power dissipation in the screen grids mounts: ongoing overdrive does overload the tube for the duration, and its lifetime is shortened.

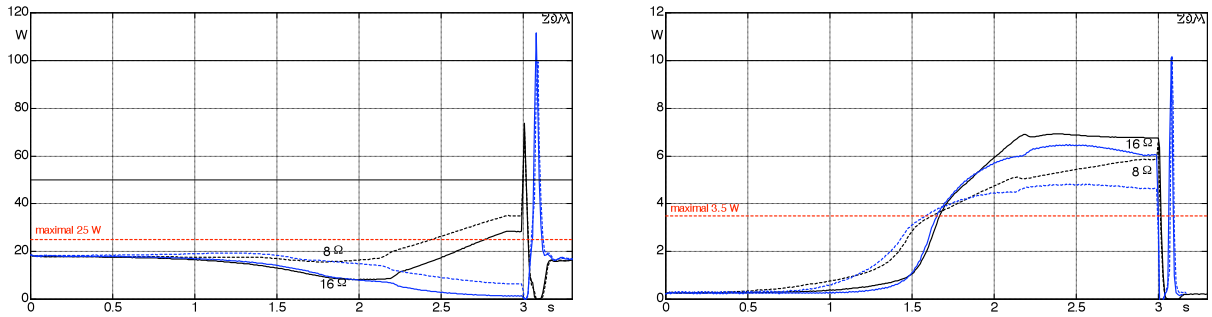

Fig. 10.5.55: JTM-45: power dissipation at the plate (left), and at the screen grid (right). From 0 to 3 s, the level of the input sine-tone rises linearly by 30 dB; at t = 1.5 s a power of P = 30W at an 8-Ω-load is reached.
At t = 3 s the drive signal is switched off; balancing processes in the capacitors of the power stage follow.
The power dissipation of one of the power tubes (TAD KT-66) is shown in black, the one of the other in blue.

The transmission characteristic from the input of the differential amplifier to the power output is shown in **Fig. 10.5.56**. As soon as the power amp is overdriven, the curve looses its point-symmetric shape, and the duty cycles change. The reasons for this are potential shifts in the differential amplifier (phase inverter) and the grid-current flowing in the power tubes. Until just short of the drive-limit, the output signal is proportional to the input signal as can be seen in the left hand picture. As overdrive occurs, the output voltage experiences limiting but also becomes increasingly asymmetric, and consequently the characteristic curve shifts (the average value needs to remain zero). Since the limited signal does now include several rather than a single frequency, phase-shifts occurring in the output transformer (acting as a high-pass) start to take an effect. The transmission characteristic is not memory-free anymore but decomposes into a rising and a falling branch. To retain sufficient clarity, Fig. 10.5.56 does not show the corresponding hysteresis-loops but average values. If a loudspeaker were to be connected rather than the ohmic load resistor, the complex impedance would result in even more complicated curves.

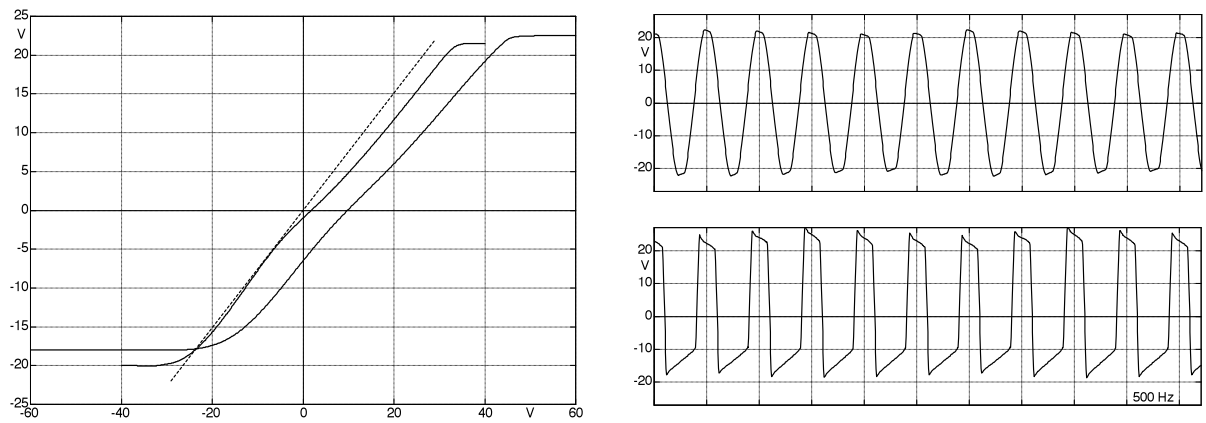

Fig. 10.5.56: Idealized transmission characteristic (left); time-function of output for ohmic nominal load (right).

The **output transformer** (OT) does influence both the frequency response and the non-linear distortion of the power stage – but there is no mystery about that. In order to conduct measurements as well as listening tests we put together a test-circuit that allows selecting between 10 different output transformers via a switch. Among others, we had in the running: a Marshall-JTM-45-transformer, a TAD MJTM45A, a Hammond 1750Q, an IGOT-JTM45 plus a few 4-kΩ-transformers. In the frequency range relevant for the guitar, all transformers showed practically the same frequency response. Regarding maximum power and harmonic distortion, there is no more than a just-about-noticeable difference between a 4-kΩ-OT und an 8-kΩ-OT. Consequently, these results cannot be the basis for advising a swapping of transformers. When comparing two OT's, the most important parameter is the transformation ratio. If two top-quality transformers produce audible differences in sound, this is most likely due to a different **transformation ratio**. However, turns-ratios can be set and checked very easily and precisely, and therefore any exorbitant pricing of a transformer is not justifiable merely on the basis of a special transformation ratio.

This is a good point to take a short side-trip into **advertising psychology**: as an exceptionally gifted transformer winder, what can you do to increase your turnover? You could write: "we are the best!" … but they all write that. Rather, you could motivate an independent trade-journalist to write an editorial contribution about, say, "Restoring Marshalls". You then find a well-known musician not happy with the sound of his/her Marshall – and off you go. Taking stock: a boring sound, odd harmonics (!), the worst Marshall since dinosaurs (of any kind) roamed the planet. After this diagnosis, on to the therapy: swap components! You will want to grab: genuine carbon-resistors, yellow or orange capacitors (depending on which supplier forks over a more generous subsidy), and of course: a new mains transformer (it supplies all that power, after all), and a new output transformer (all that music needs to pass through it), and, since we're at it, throw in a new choke. Ah – now we're in brown-sound-city: the best Marshall ever heard! Last, make the well-known musician rave about the unbelievable improvement in sound, and make him/her recommend that everybody installs these wicked transformers. Now, it only remains to hope that nobody checks *www.tone-lizard.com/marshall-myths*, where a discussion can be found mentioning – with relish – that for repairs frequently a damaged Marshall-transformer was exchanged for a low-cost no-name transformer … and *not a single complaint* was ever received. Delightful stuff.

It is normal and necessary that manufacturer advertise their products; that they hire musicians to praise the unrivalled sound may be criticized but there's not much that can be done about that. From a technical point of view, nothing stands against swapping a correctly working Marshall transformer for an expensive clone. It is easily conceivable that a guitarist feels better after the swap than before – but that has different reasons then.

The JTM-45 and its output transformer have achieved cult-status. Supply (meager) and demand (high) now regulate the price (enormous). In Doyle's Marshall-book we read, however, that the differences to the 5F6-A-Bassman are in essence due to the different loudspeakers (Celestion 12" vs. Jensen 10"), the different input tube (12AX7 vs. 12AY7), and the higher negative feedback in the power-stage of the JTM-45. The RS-output-transformer is not the reason for a special sound, as Doyle cites the design-director of Marshall. Hopefully, nobody still believes that a steel-chassis will make the amp sound different compared to an aluminum-chassis. You over there still do? Be informed that this is another myth. Aluminum has paramagnetic characteristics while steel is ferromagnetic?! So? The effects on the sound are about as dramatic as the color of the control-knobs is.

There is, however, a sound-determining parameter that has so far been investigated too little: the **2nd-order harmonic distortion**. In a transistor amplifier we usually pay close attention to symmetry, and consequently even in overdrive mode the 2nd-order distortion is reduced to insignificant levels. The tube power-stage, on the other hand, shows a rather different behavior (e.g. Fig. 10.5.56): as the overdrive increases, the duty-cycle changes and k_2 may not be neglected anymore. **Fig. 10.5.57** shows distortion measurements: the differences between the individual curves are rather substantial. What is the reason? These are different tubes (all KT-66). TungSol, TAD, and several original GEC-KT66 from the good old days. They are accredited with qualities that allegedly are not achievable anymore today, and so a pair of GEC-KT-66 may be offered (at the time of writing) for 280 Euro. That could be $699, as well, if we jump to the other side of the Atlantic (Ebay, December 2013). Stiff prices, indeed.

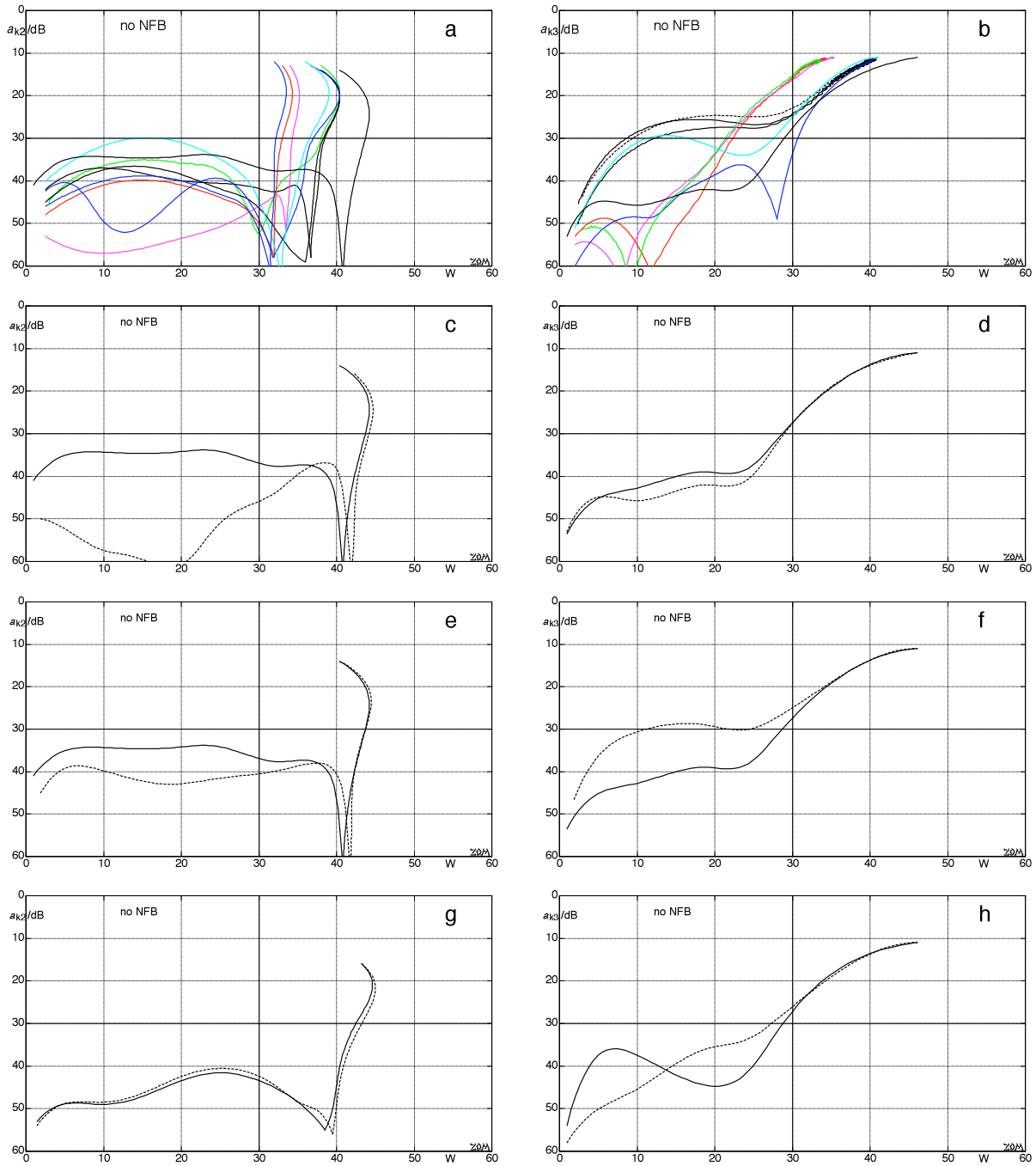

Fig. 10.5.57: 2nd order (left) and 3rd order harmonic distortion, KT-66, R_{aa} = 8 kΩ, 8-Ω-load at the 8-Ω-output.

The first line in **Fig. 10.5.57** shows the scatter-width across 8 different KT-66-pairs. For pairs **c and d**, only one plate-resistor was changed: 82kΩ/108kΩ were included rather than the usual 82kΩ/100kΩ (differential amplifier). Immediately, the power-stage anti-symmetry changes, as does the 2nd-order harmonic distortion. For **e and f**, the offset voltage at the grid of the power tubes was changed from -53V/-50V (13mA/13mA) to -48V/-48V (25mA/19mA). This KT-66-pair was 'matched' only to a rather lukewarm degree, despite a 4-digit-coincidence of the numbers on the sticker. Increasing the plate-current increases the 3rd-order distortion; the asymmetry in the current to some extent compensates for the disparity in the tubes (a_{k2}). For **g and h** another KT-66-pair was used, and the bias-current was changed from 13mA/13 mA to 17mA/17 mA. It is surprising that the KT-66 requires such a small bias-current for low distortion (it was operated with R_{g2} = 1.5 kΩ for these measurements).

The bias-current could be easily adjusted in this JTM-45; no potentiometer is, however, foreseen to set the symmetry. In fact, there are two values of relevance here: differences in the drive circuit (plate-resistors in the differential amplifier), and the offset-voltages at the grids of the power tubes. Of course, the respective individual KT-66 adds in, as well. Frequently, carbon film resistors are recommended in order to achieve the original sound. However, such resistors normally have a tolerance of ±10%! The variation of a plate resistor from 100 kΩ to 108 kΩ is comfortably covered by this tolerance span, but it will change k_2 (at c) by more than a factor of 10! If such variations are indeed considered to be relevant, it is not necessary to shell out more than 15.000.- Euro for an old JTM-45 – one or two potentiometers added into the circuit of a reissue amp will do fine (at a price of 3 Euro per piece).

The notion that KT-66's produced today will not match the original data holds, in this experiment, only for the TAD-tubes (which in the meantime may well be supplied by another manufacturer – we did not investigate this aspect). The TungSol-KT-66's are not generally worse than the original GEC-tubes – quite the contrary. We had 8 GEC-tubes at our disposal for the measurements. True vintage! Correspondingly, they were handled with great care. One of these tubes was practically useless, two others had a gain so low that they could not be used. The remaining KT-66's worked well but their data corresponded only very moderately, despite the "7500/7500"-marking on one pair and the "6500/6500"-marking on another pair. The seemingly 4-digit correspondence ("matching") did not keep the tubes from featuring different gain – which will influence the harmonic distortion.

The above observations warrant the warning not to acquire vintage tubes (so-called NOS) from unknown sources. This especially holds if the prices are significantly higher than those of new tubes. At present, a pair of KT-66 is about 65 Euro, and it would be unwise to pay much more. While it is indeed possible that vintage tubes on the market are well paired and have been used little, and also feature small grid-currents and a good vacuum, they may just as well be of abysmal quality. What can you do if (prepaid with an enormous sum) a parcel arrives from far-away lands with tubes for which merely the label is correct? It may be rather profitable to re-sell a cosmetically perfect replica (bought for a few Yuan) for $699 per pair – but let's mention that only in passing. It does of course not imply, that all NOS-tubes necessarily are fakes (for more information search the web for "faked tubes").

To avoid that the results shown in Fig. 10.5.57 are interpreted as stellar peculiarity of the JTM-45: please take note of another warning. Similar curves are to be found with Fender power stages, as well – this is not exclusive to Marshall! The plate resistors in the differential amplifier, the degree of pairing of the power tubes, the bias current, the negative feedback – all this determines the behavior of the power stage. The equation "vintage = great" does not compute!

Fender – VOX – Marshall, the holy trinity: sure, it will not command the respect of every guitarist, but the constantly recurring chorus in the "vintage"-columns of magazines has generated the widely held opinion that the primordial VOX (or the proto-JTM-45, or the ancient Bassman) is unequalled sonically, and easily justifies the $10.000 or 20.000 asked today for the old originals. And of course, it is alluring to elect the top dog of this troika: "compared to the 1959 Fender Bassman it is modeled after, the JTM outperforms its alter ego with ease" (Gitarre&Bass, 7/2006). Onto the podium – long live the myth.

How should we imagine the scenery at the beginning of the 1960's, when this legendary amp came to life? Maybe like this (as some would have it): 39-year-old Jim sits behind his drum kit, squeezing some sophisticated triplets between the screaming guitar-arpeggios, and thinks *'that doesn't sound like Hardrock at all – I'm gonna build 'em a new amp with the right brit-brown sound'*. And then he tells Ken: *'get on with it'* – and the result is the JTM-45 with its unparalleled distortion sound? Maybe it was like that, with Jim the Rocker? This image does not really fit the picture found in the books about Marshall: a friendly gentleman sporting suit and bow-tie who probably makes his sticks dance across the skins in a more gentle manner. Wikipedia sees the start of the **Hardrock**-era in 1969 but not in 1962. We know that Ritchie Blackmore, Jimi Hendrix, Pete Townshend and many others came to fame using Marshall amps, and it is easily imaginable that they voiced requests for more power – could that have been in 1962, though? Townshend played (according to Wikipedia) in a Dixieland-band in 1959, then graduated to Skiffle, and the Who gets off the ground only as late as 1964. Deep Purple forms in 1968, Hendrix starts his Experience in 1966, and Clapton plays with the Yardbirds in 1963, miles away from any Beano-like tone. It is also sufficiently well documented that Brian Poole (with his Tremeloes) was not an early exponent of Hardrock.

No contest: Jim Marshall has deservedly earned his medal as amp-pioneer – summa cum laude, without any doubt. That does not imply, however, that the JTM-45 was developed and optimized as distortion-heavy amp, even if this rumor is circulated within fan circles. Folks, read closely what Ken Bran states in the Marshall book: *"It was a bass amp we originally wanted ... but the guitar sound was too good to pass up."* The differences existing between the bass, guitar, organ and PA-variants of the early Marshalls are limited to two small bridging-capacitors to boost the treble. Had the JTM-45-circuit been developed to generate special distortion, it would have also distorted vocals amplified by the PA-version – except for the different treble gain, all these amps were identical. Many guitarists found (and continue to do so) that the JTM-45 sounds really good when overdriven, but already in the description of the distorted sound we find differences: according to Wikipedia, the Bluesbreaker combo (Model No. 1962) was the amp that *"first led to the breakthrough of the typical Marshall sound"*. However, in Gitarre&Bass (07/2006) we find the statement that this same amp produces *"never a Marshall-typical distortion sound"*. The author, writing a monthly column about vintage amplifiers, is somewhat of a Nostradamus-of-the-tube-amp (i.e. not looking into the future but backwards-oriented – we are talking *vintage* here!), and in terms of interpretation simply congenial. A sample: *"and the result (JTM-45) differed, in the end, strongly from a Fender Bassman"* (G&B 07/2006) **versus** *" the first of the so-called JTM-models were therefore rather authentic copies (of the Fender Bassman)"* (G&B 2/2005). Just like with Nostradamus: it all depends on the year. Not a problem for anybody bred and raised in Munich, Bavaria, and familiar with local poet Karl Valentin who wrote: *"it has expertly been calculated that the Lake of Starnberg* (a well-known beautiful lake south of Munich) *is, at the same time, deep, shallow, long, short, narrow, and wide.*

Fender Super-Reverb

A typical medium-power Fender amplifier holds two 6L6-GC's, and the Super-Reverb is a good example. The cathodes of the power tubes are connected directly to ground, and a separate diode generates the negative offset-voltage: no doubt at all – this is textbook-class-AB-operation. Ahead of the power tubes we find a differential amplifier, following them the output transformer with a connection to the negative feedback loop. All in all it is a model for the way Fender power stages looked like in the 1960's. Still, there are individual idiosycracies: the driver-tube may change (12AT7 instead of 7025), the coupling capacitors, too; small blocking capacitors are discarded, then they return again – and even a "Presence"-control is found in the 'Super' for a short time. The Super-Reverb investigated in the following has the AB-763-circuit originating in the 'Blackface-era' i.e. in the golden 1960's.

Fig. 10.5.59 depicts the output characteristics for ohmic loading of the 8-Ω-output (this specimen of the amp had a transformer with such a connection installed). For the specified load, the "knee" of the 0-V-curve is almost exactly met, indicating an optimum transformer dimensioning. As the drive level rises, the curve is shifted towards smaller voltages.

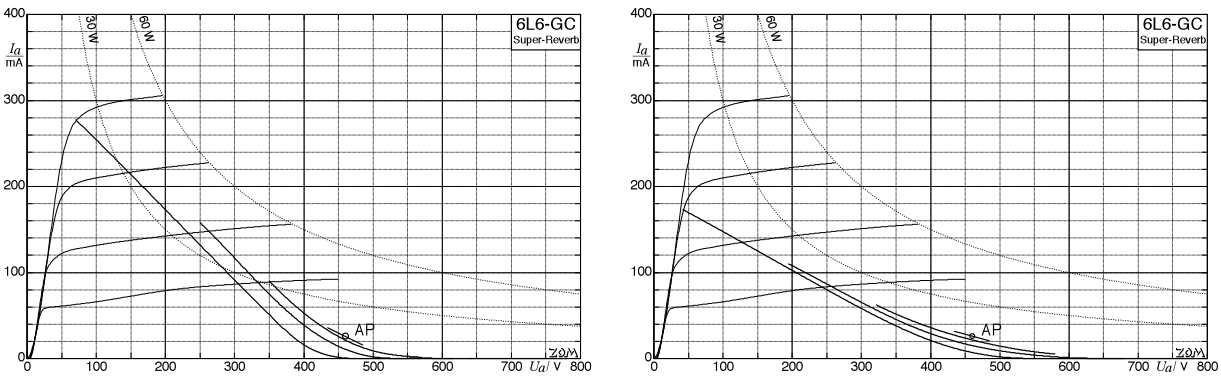

Fig. 10.5.59: Characteristics for ohmic load of 8Ω (left) and 16Ω (right).
Note: the output transformer used here also had an 8-Ω-output on top of the regular 2-Ω-output.

The negative feedback in this power stage is not as strong as it is in the JTM-45, and therefore the loudspeaker impedance is more clearly represented in the transmission frequency response (**Fig. 10.5.60**). For all these diagrams, it is important to recognize that the exact shape of the curve depends on the specific loudspeaker: the loudspeaker resonance, which is about 75 Hz in the given example, may rise to over 100 Hz with other speakers. This of course has an effect on the sound (compare to Chapter 11). If the power stage is overdriven, the influence of the speaker diminishes and the characteristic becomes closer to that of a voltage source. This is shown in upper curve of the left-hand picture.

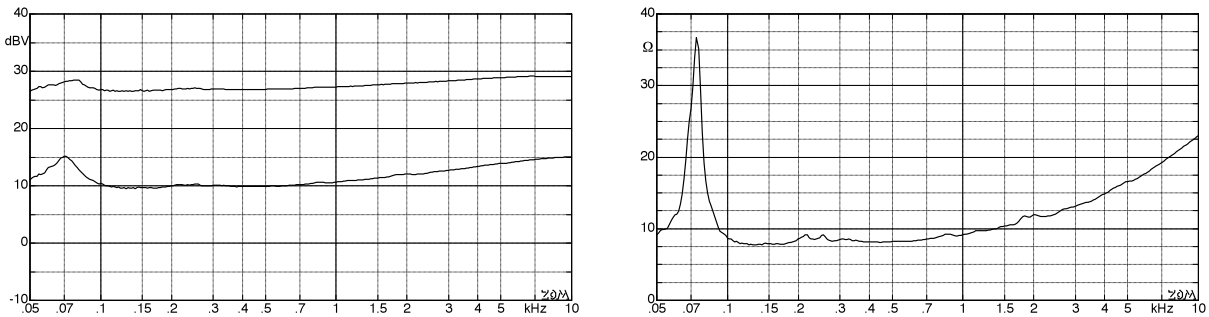

Fig. 10.5.60: Frequency response of a Super-Reverb power amp, 8-Ω-output loaded with 4xP10R (left).
Right: magnitude of the loudspeaker impedance (4xP10R, cabinet set up in reflecting surroundings).

The transmission characteristic of the power stage for 8-Ω-loading is shown in **Fig. 10.5.61**. As the drive level rises, the curve decomposes into two branches that slide apart. As has already been noted, the reason is the polarization of the coupling capacitors.

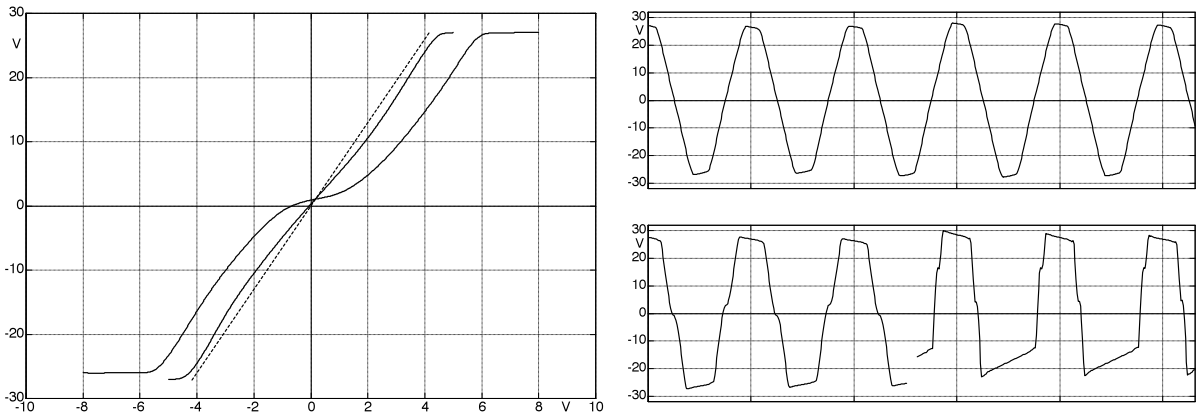

Fig. 10.5.61: Idealized characteristic (left), output time-function with ohmic nominal load (right).

We can see from **Fig. 10.5.62** that the output power of 40 W (as specified e.g. in the 1968 catalog) is actually achieved. This is in sharp contrast to the JTM-45, the replica of which is advertised by TAD (in 2008) with "about 45 Watt" but reaches merely 30 W. The minimum of the harmonic distortion is due to the progressively curved characteristic that changes the direction at the onset of distortion. The strain on the power tubes is similar to the JTM-45: the screen grid is overloaded for overdrive operation with a high-impedance load (**Fig. 10.5.63**). One significant difference is found in the input capacitor: if it is only 1 nF (AB763), the plate is overloaded less (compare to Fig. 10.5.55). There are, however, also Fender amplifiers with a larger input capacitor (e.g. 10 nF).

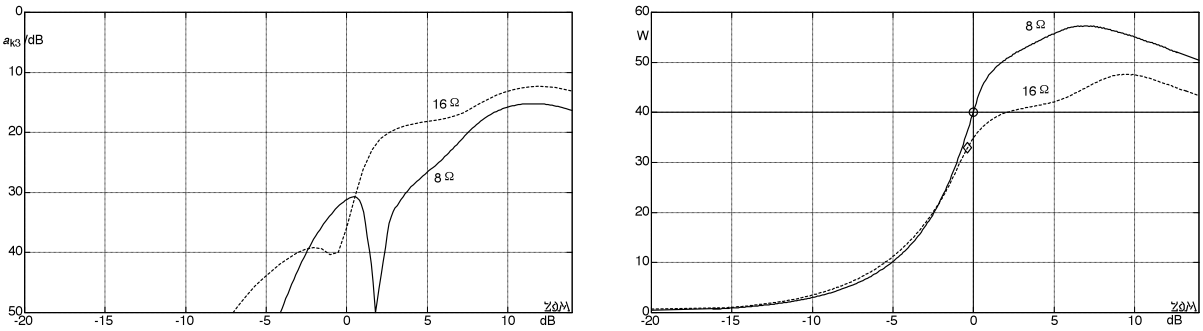

Abb. 10.5.62: Super-Reverb: harmonic distortion, output power at the 8-Ω- output with 8-Ω- and. 16-Ω-load.

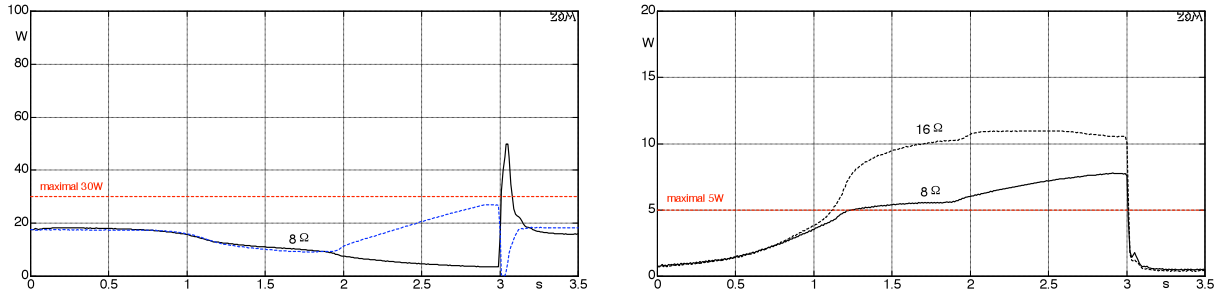

Fig. 10.5.63: Power dissipation at the plate for both output tubes (left); power dissipation at the screen grid for two different load impedances (right). The level of the input signal (500 Hz) rises linearly from 0 – 3 s, switch-off occurs at t = 3 s. From t = 1.3 s, the power stage is overdriven.

Comparison of power stages

The 1960's holy trinity: VOX, Marshall, Fender. Of course there are also Gibson, Ampeg, Hiwatt and many more, but the 'big three' stand out. So, what makes for the difference between these amplifiers or, rather, between the respective power stages (to do this chapter adequate justice)? This question cannot be answered generally because there is not *the one* Marshall- or Fender-amp. Even at VOX, the AC-30 ran through several production variants. For Fender, Dave Funk lists 250 pages of schematics – and still has not captured all Fender amps. Practically every amplifier model (e.g. the Bassman) was, over the years, built in many variants, and there are many models to begin with. It is therefore impossible to speak of one Fender-typical circuit, or of one Fender-typical sound. The situation is similar for Marshall – only the AC-30 remains reasonably true to itself, although even here there are modifications, e.g. the models developed for the US that only seemingly were similar to the UK-standard.

Even when concentrating on only three special power stages, a comparison turns out to be difficult due to many small differences in detail. Most important are category of output-power, negative feedback (and correspondingly the internal impedance), and balancing processes during overdrive conditions. Even the loudspeaker needs to be considered although it is not part of the power stage: its impedance determines the load on the power stage and thus the frequency response of the latter. The circuitry preceding the power stage plays a considerable role, as well: is it of high or low impedance, and what voltage can it offer without distorting? If the power stage were a linear and time-invariant system, we could record its frequency response and have a good starting point for comparisons. However, guitar amps are subject to overdrive (i.e. they are operating as non-linear systems), and therefore a small-signal analysis allows for only very limited conclusions on their behavior.

To illustrate the problems appearing when comparing amplifiers, let us look at the VOX AC-30 and the Fender Super-Reverb. The VOX offers 30 W, the Fender 45 W. In the VOX-cabinet we find two 12"-**loudspeakers** while four 10"-speakers are deployed in the Super-Reverb. If we allow for each power stage to work with its original speakers, we not only compare the power stages but also the loudspeakers. Should we consider connecting the VOX-speakers to the Fender, we risk blowing them because Celestion specifies only a 15-W-load for each speaker. Moreover, the nominal impedance the Super-Reverb is specified for is 2 Ω, while it is 16 Ω for the VOX. One could re-solder the VOX-speakers to a 4-Ω-configuration, but that would result in yet another different scenario. How about the other way round: operating the VOX with the Fender speakers? That would work in terms of power capacity, but the issue with the different output power remains: it could result in differing loudspeaker distortion (with the sub-harmonics being level-dependent).

Therefore, the chosen approach would have to be to use only one and the same loudspeaker for all amps to be compared. What would remain now as power-stage specific differences? First, the **internal impedance**: it is high in the VOX, medium in the Fender and low in the Marshall. The speaker impedance will therefore more or less shape the frequency response. At resonance, the loudspeaker impedance can rise to 40 Ω or even 150 Ω, implying a voltage-level difference of almost 12 dB for a high-impedance source and an almost unchanged level or a low-impedance source. This is an enormous difference that is neither due to the power stage by itself nor caused by the loudspeaker by itself (**Fig. 10.5.64**). Even though the power stages are neither pure voltage sources nor pure current sources, the corresponding difference between an AC-30 ($R_i \approx 80\ \Omega$) and a JTM-45 ($R_i \approx 2\ \Omega$) is considerable.

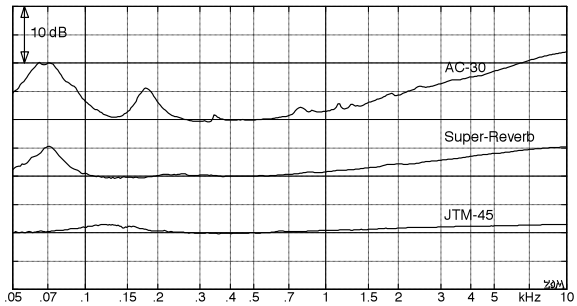

Fig. 10.5.64: Frequency responses from phase-inverter input to loudspeaker output.

AC-30 with 2x12"-Celestion in combo-enclosure,
Super-Reverb with 4x10"-Jensen in combo-enclosure,
JTM-45 with 4x12"-Celestion in separate 1960AX enclosure.

The reason for the different internal impedances is the **negative feedback** (NFB) **in the power stages**: it is strong in the JTM-45, somewhat less strong in the Super-Reverb, and non-existent in the AC-30. Besides influencing the internal impedance, the NFB also has an effect on the non-linear distortion of the power stage: this distortion is stronger in the AC-30 and smaller in the Super-Reverb and, in particular, the JTM-45. The type of distortion varies, as well: with increasing overdrive, the duty cycles in the JTM-45 and the Super-Reverb change, and correspondingly 2nd-order **distortion** mounts. Conversely, the output signal remains largely half-wave anti-symmetric in the AC-30, with k_3 remaining dominant.

The **output transformer** influences the output signal, too – though less than first expected (see Chapter 10.6.5). In the low-frequency range, harmonic distortion caused by the transformer can become audible – but only for really low-quality transformers. All transformers investigated here gave no cause for complaint. Because not all transformers have the same turns-ratio, the frequency responses differ a little; this, however, is no secret science – in essence this is a matter of the number of turns in the windings.

The ratio of **impulse power** to **continuous power**, and the **hum-interference-modulation** is not alone a characteristic of the power stage but the power supply is involved, as well. The Super-Amp 5F4 had a capacitor of 16 µF connected after the rectifier tube, and another 16 µF after the choke. That was indeed rather modest, so the successor receives 40 µF / 20 µF. With Marshall, the JTM-45 first included 32 µF / 32 µF, but the model 1987 filtered with an ample 100 µF / 50 µF. Started out with 16 µF / 16 µF, the AC-30 was upgraded to 32 µF / 32 µF later. It is a well-known fact that all these electrolytic capacitors often had considerable tolerances.

The plate resistors in the **phase inverter** are a science in themselves: we have 82k/100k with the 7025 in the 6G4, 100k/100k with the 12AT7 in the AA763, 47k/47k with the 12AT7 in the AB568. In the Marshall, an ECC83 with 82k/100k is at work, and in the VOX an ECC83 with 100k/100k. The anti-symmetry of the phase-inverter outputs influences the even-order distortions in the power stage. The scatter of component values can have an extremely strong effect, and, of course, the equality of the **power tubes** plays a role, as well ("matching").

The power tubes: EL84, 6V6GT, 6L6GT, KT66, EL34, KT88, and relations. This is a difficult topic because there is not "the" 6L6GT – the scatter can be very wide. The acquisition of a large number of 6L6GT (say 12 pieces) does not help here, either: if all twelve tubes are from the same production batch, they might have similar parameters, but if we later buy another pair, the parameters might well be entirely different. It has already been elaborated that "selecting" and "matching" are no cure-alls, either (Chapter 10.5.11). The measurement results listed in the following are therefore to be taken merely as a snapshot to provide orientation values of limited general validity.

10.5.13 Comparison of power tubes

What happens if the two 6L6GC in a Vibroverb are swapped for a pair of KT66 – or for a pair of EL34? Citations of how these tubes allegedly sound have already been given on the preceding pages. Let us first disregard the sound – how do the electrical data change? First, there is the **heating**: two 6L6GC require a filament current of 1.8 A, two KT-66 demand 2.6 A, and two EL34 already push this to 3 A. The mains transformer is therefore put under different strain, but let us insinuate by all means that it can take the additional load for the short term. The bias voltage at the grids (i.e. the bias current) needs to be adapted, of course – and now what? Does the frequency response change substantially due to the tube-swap? What about the harmonic distortion? More generally: which part do the tubes play in the operating behavior of the power stage?

The simple solution: it is the output power that depends on the power tubes – and that's it. You may or should add a few bits here and there, but in essence, this is the sobering answer. We do find differences already with regular instrumentation, but the relevance to the sound remains very modest. The measurements discussed in the following were taken from a Marshall power stage that was, however, operated via a stabilized 400-V-power-supply. One of the two plate resistors of the differential amplifier was adjustable in order to balance out different gain of the power tubes. The primary impedance of the output transformer was R_{aa} = 3.5 kΩ; the resistors at the screen grid had 1.5 kΩ each. To emphasize any differences, the negative feedback in the power stage was deactivated. **Nominal load** implies that an 8-Ω-resistor (purely ohmic) was connected to the 8-Ω-output. The signal generator was directly connected to the input of the differential amplifier (→ v_U).

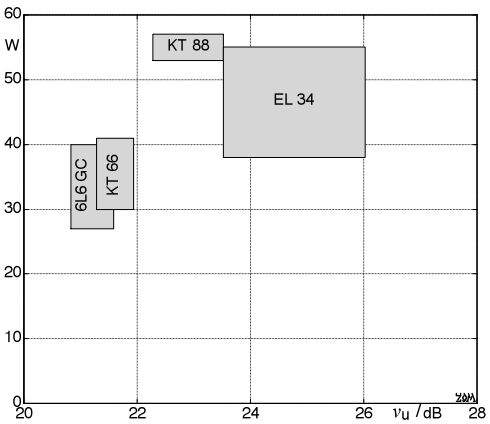

Fig. 10.5.65: Output power vs. voltage gain.
6L6GC from GEC, TAD, JJ, Ultron, TungSol, Sovtek;
5881 from Sovtec, TungSol;
KT66 from TAD, TungSol, Marconi;
EL34 from TubeTown, TAD, JJ, EH, Valvo;
KT88 (and 6550) from Sovtek, GEC, EH, SED;

Power measurements were taken at 500 Hz at the onset of clipping and with nominal load. Any influences due to the power supply were eliminated using a stabilized plate-voltage (400 V)

Fig. 10.5.65 gives an orientation regarding the output power and the gain of the power stage with different tubes. The sample was very small (20 6L6GC, 10 EL34, 10 KT66, 8 KT88), and therefore it is to be expected that the market will offer specimen with data lying outside of the grey areas. Already the tubes measured here show considerable scatter in the maximum power: a fresh pair of 6L6GC may yield the datasheet-conform 40 W, or a meager 27 W. For the EL34, the span extends form 38 W to 55 W, and consequently the changeover 6L6GC → EL34 could bluntly double the power … or reduce it some. In any case, the probability that the voltage gain goes up by about 2 – 5 dB is high. The higher maximum power could lead to stronger distortion in the connected loudspeaker – but this should not tempt us to generally attest more distortion to the EL34. If at all, these would be indirect tube characteristics. How much the power stage itself distorts, that will be subject to the following analyses.

All these analyses were done at **500 Hz and with nominal load**. Before we investigate the harmonic distortion, let us take a look at the **transmission characteristic** i.e. the mapping of the generator voltage onto the output voltage generated across the 8-Ω-load-resistor. With a small bias-current (20 mA), we see a saddle-point for small drive levels, in other words a progressive curvature of the characteristic. This changes into a degressive curvature with high bias-current (60 mA). All three curvatures are depicted in **Fig.10.5.66** in an idealized manner.

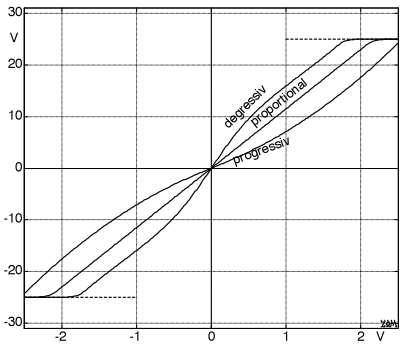

Fig. 10.5.66: Transfer characteristics

The progressive characteristic rises from the origin with increasing slope, the proportional dependency shows a constant slope, and for the degressive curve the slope decreases with increasing drive. At 25 V the so-called clipping (limiting of the ordinate values) sets in. These curves are, however, idealized; for the real tube there is no perfect proportionality: all curves are "somehow bent". The exact shape depends on the geometry of the tube-electrodes and on how equal or unequal the two power tubes are; it will therefore be different for each push-pull power stage. See Chapter 10.5.3 for the basics of push-pull operation.

From the idealization on to real tubes: **Fig. 10.5.67** shows three transmission characteristics of a Sovtek 5881♣. It corresponds best to the above idealization (which does not necessarily hold for all tubes of this type, and much less for Sovtek in general). Seeking the least distortion, we would have to choose the middle curve (40 mA). The Groove-Tubes 6L6GC shown next also allows for a good proportionality, although it requires 60 mA cathode-current, which makes – at 400 plate-voltage – already for a pretty hot operation. The Tung-Sol 5881 again is more similar to the Sovtek 5881 – so much so that the conjecture finds support that the two tubes may differ only in the labeling. The next tube, an old Marconi, can keep up well with the others although this is not a 6L6GC but a KT-66.

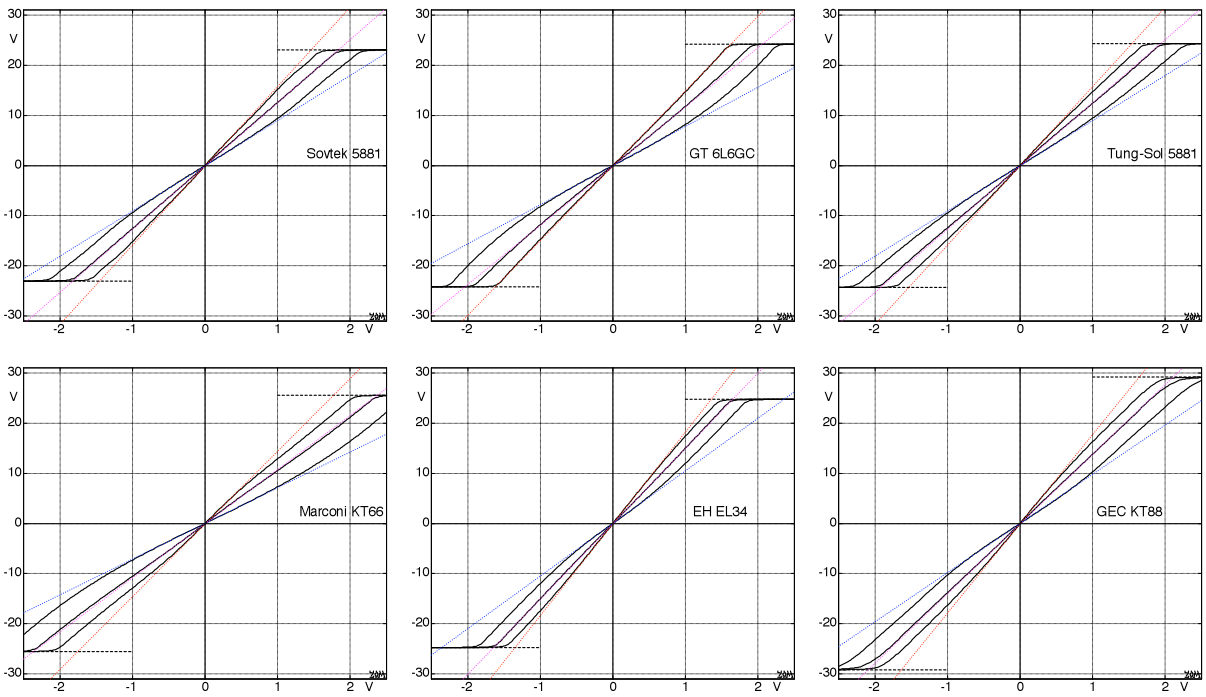

Fig. 10.5.67: Measured transmission characteristics, nominal load, I_K = 20, 40, 60 mA.

♣ The 5881 is the professional variant of the 6L6GC.

For the next tube, the family of characteristic curves looks similar, too, as it does for the last specimen in this overview. These are entirely different tubes, however: here we have a pair of EL34's and a pair of KT-88's. We do see some differences at the drive limit but the basic curves are very similar indeed. For these 6 pairs, that is! The tubes from **Fig. 10.5.68** show that more strongly bent curves exist, as well – to a varied degree. For the JJ-6L6GC we see a pronounced ripple while the TAD has a more degressive characteristic – this shows that the difference between a 6L6GC and a KT-66 is not necessarily bigger than the difference between two KT-66's.

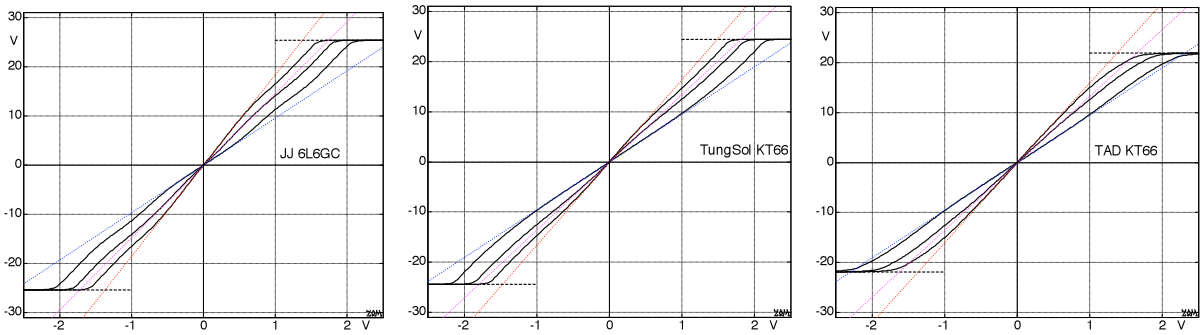

Fig. 10.5.68: Measured transmission characteristics; nominal load, I_K = 20, 40, 60 mA.

The x-vs.-y-depiction is not very suitable to clarify how well the transmission characteristics can follow the ideal proportionality – the **analysis of the harmonic distortion** delivers better results here. The level of a 500-Hz-tone was increased by 30 dB over the course of 4 seconds, and at the same time the levels of the first 20 harmonics were extracted from the output signal of the power stage (software Cortex VIPER). The results are shown in **Fig. 10.5.69**.

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Fig. 10.5.69: 3rd-order harmonic distortion vs. output power, 500 Hz, nominal load, NFB deactivated.
These figures are reserved for the printed version of this book.

In Fig. 10.5.69, a maximum of the distortion attenuation (= distortion minimum) appears around 40 W. This is an effect of the change in curvature in the range of the clipping-onset; the 3rd harmonic changes its algebraic sign here. As different the curves may look – the non-linear distortion can be easily reduced to inaudible levels by choosing the appropriate bias-current. However, "appropriate" means a whopping 60 mA for the GT-6L6GC, but no more than 30 mA for the JJ-6L6GC. In the case that distortion is heard when comparing power tubes: that may simply be due to an inappropriate bias-current!

We have all heard or read opinions related to the 'tube-sound': *"simply unplug the two 6L6GC and plug in two KT-66 – you gotta hear that difference!"* Just like that, without considering the bias-current? It is almost certain that, with such a makeshift setup, differing modes of operation are evaluated rather than the difference in the tubes per se. Setting the bias-current via the bias-voltage at the grids will not remove the problem: mind you, for the same bias-voltage at the grids, three "premium matched" 6L6GC-pairs show sizeable variance in the bias-current (30 mA vs. 40 mA). The voltage gains for the two half-waves did not match, either♣, with 10% difference. This problem persists in particular with the famed NOS-tubes because normally there will not be 20 of them available to choose the ones matching best.

Fig. 10.5.69 compared four beam-tetrodes. Does the EL34, a real pentode, show other curves? Yes and no, as we see from **Fig. 10.5.70**. In the details, there are strong differences and in particular there is more power, but we cannot speak of a generally different behavior. If we take k = 3% as the limit for audible non-linear distortion in a guitar amp, all tubes generate audible distortion only just before going into clipping, if the bias-current is set correctly. They also show the same type of increase of the THD. Moreover, we must not forget that the THD will decrease significantly as we (re-) activate the negative feedback.

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Abb. 10.5.70: As Fig. 10.5.69, but for EL34 and KT-88; without NFB, 20 – 60 mA.
These figures are reserved for the printed version of this book.

♣ Amongst tube retailers, the readiness to match and pair (up) does not seem to be particularly distinct ...

Time for an **interim statement**: yes, the tubes under investigation show variances, but in essence this is limited to power output and gain. While we find differences between 6L6GC and EL34 regarding the individual distortion characteristics, similar differences are observed between two pairs of 6L6GC. Is it, however, sufficient to analyze only these parameters? What about the **internal impedance** (also termed source impedance)? **Fig 10.5.71** shows the corresponding measurements. Pentodes are of high impedance, the frequency dependency mainly stems for the output transformer (Marshall). The resonance of the winding receives differing dampening from the power tubes; this makes for a different height in the maxima. The left-hand picture depicts the impedance curves for six pairs of 6L6GC (lines), two pairs of 5881 (dashed), and one 6L6WGC-pair (dotted). The right-hand picture shows the results for two KT-66-pairs and for three pairs of EL34. What's interesting: the lines close to each other are the dashed line (KT-66) and solid line (EL34) i.e. they are not the ones that would "belong" together. So again, we do not see a general difference.

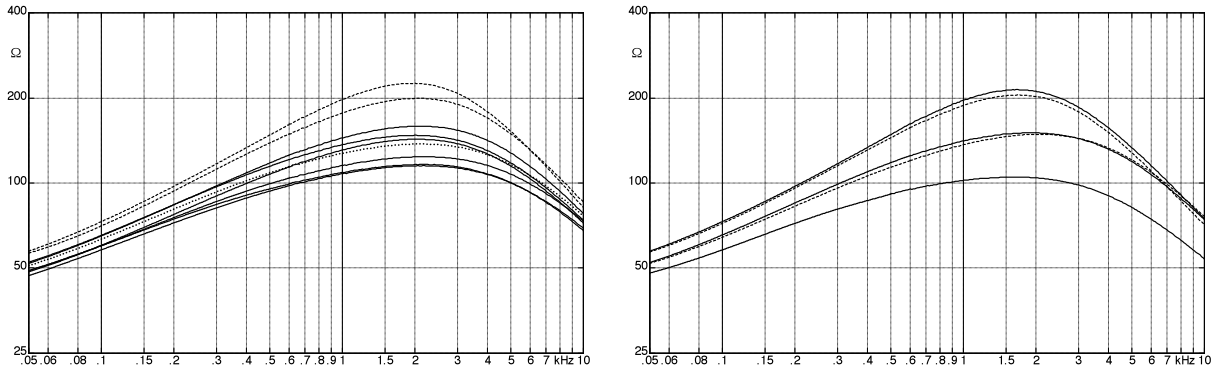

Fig. 10.5.71: Frequency response of the internal impedance measured at the 8-Ω-output; I_K = 40 mA.

The internal impedance (which is also dependent on the bias-current) influences the dampening of the loudspeaker and thus the transmission frequency response. This however holds mainly for power stages without negative feedback. Chapter 10.5.14 elucidates how much this effect looses its significance as soon as the negative feedback is in action.

Was that it? No! It is a widespread error to limit testing of power amplifiers to merely the nominal impedance as a load. Loudspeaker impedances are frequency-dependent♣, and therefore supplemental measurements with a **load of 32 Ω** follow here. The main difference occurs between beam-tetrode and pentode (**Fig. 10.5.72**): with a true pentode (e.g. the EL34), the current through the screen grid increases within the distribution area towards much higher values. This conversely implies a reduction of the plate-current i.e. a pronounced sharp bend.

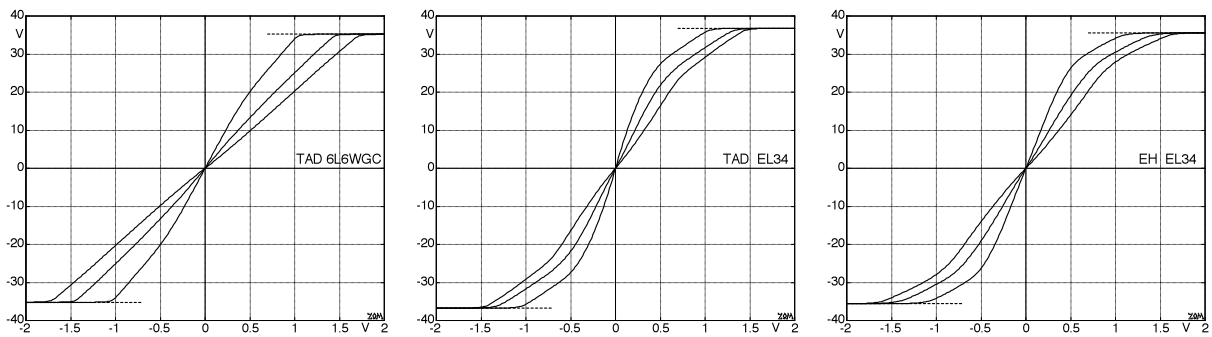

Fig. 10.5.72: Measured transfer characteristics, 32-Ω-load at the 8-Ω-output, I_K = 10, 30, 50 mA.

♣ It was already documented in Fig. 10.5.28 that the speaker does not have of a real but a complex characteristic.

The sharp bend appearing in the EL34-characteristic (Fig. 10.5.72) generates a different curve of the harmonic distortion – see **Fig. 10.5.73**. Still, we again need to heed here that the negative feedback is deactivated. As typical NFB is brought in, this effect looses its significance, as can be seen from the dashed lines.

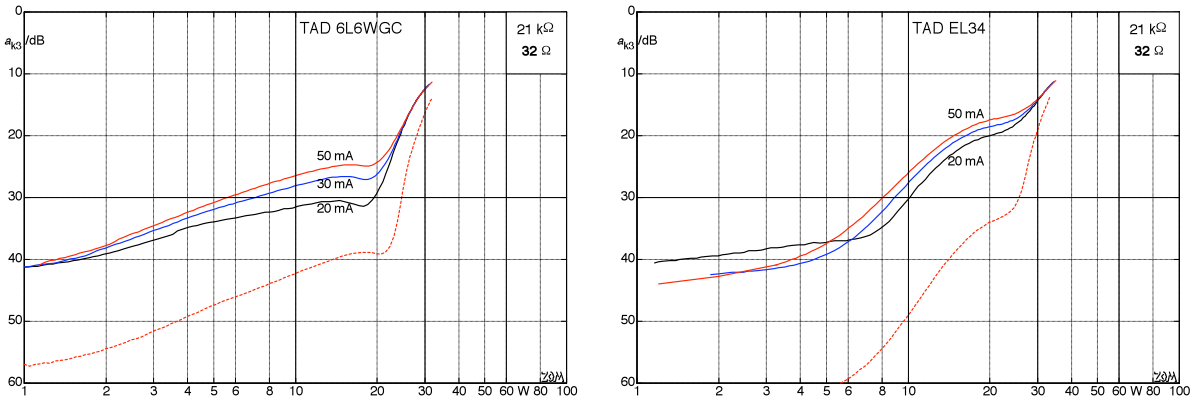

Fig. 10.5.73: 3rd order distortion attenuation vs. output power, 500 Hz, 32 Ω at the 8-Ω-output.
Negative feedback activated (dashed) and deactivated (solid line).

As different as the individual distortion attenuations might be: if we take 30 dB as audible limit, only maximum power and gain are left as main criteria (with the NFB activated). Because the EL34 has (in the present series of measurements) 2 – 5 dB more gain than the 6L6GC, the frequency response of the amp with active NFB changes somewhat. The higher the gain (or the transconductance) is, the stronger the effect of the NFB, and the smaller the influence of the loudspeaker impedance on the frequency response (**Fig.10.5.74**). In case you regard these differences as essential: simply change the negative-feedback circuit ☺.

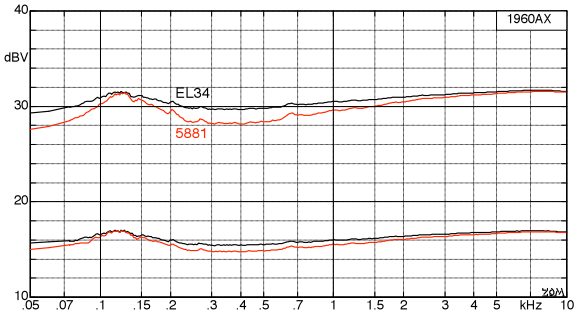

Fig. 10.5.74: Transmission frequency response.
Power stage with negative feedback, 16-Ω-output,
Voltage level at the Marshall-Box (1960-AX).
Sovtek 5881 (red), TAD El34 (black).
For the upper two curves, the power stage is overdriven.

Conclusion: a representative comparison is not possible because even the data of selected tubes include a scatter, and because the vendors do not guarantee any limit values. The sample analyzed above shows measurable differences between various 6L6GC, 5881 (6L6WGC) und KT66, but these will most probably lie within the assumed production scatter. They are also of secondary importance in everyday studio- and stage-operation. Within the sample, the EL34's distinguish themselves regarding maximum power and transconductance, and from this a marginal difference in the frequency response results. No investigation could be carried out regarding the **lifetime**. For example, in order to check the 10.000 h propagated by MOV, 14 months would be needed! If we would set up 15 power stages per tube-type in order to meet the minimum statistical requirements, the cost for mains power alone would amount to 20.000 Euro – that is not reasonable. Who can guarantee that after the conclusion of the test, the tube vendor will not have "his" special 6L6-WGC-STR-XXL-premium-selected built by another manufacturer? Due to the even better offered quality, as the vendor writes … or rather because he did not buy a sufficient quantity off the first manufacturer…

For the "small" power tubes **6V6GT and EL84**, another degree of freedom is added: besides amplifiers with a fixed offset-voltage at the grid (e.g. the Deluxe reverb), there are also amps with a cathode resistor (e.g. the AC-15). This resistor has several effects: the DC current flowing through it generates the offset-voltage, the power dissipated in it is lacking in the loudspeaker, and in overdrive mode it changes the operating point.

In the AC15, the **cathode-resistor** has a value of 130 Ω generating about -10 V offset voltage, and somewhat more than 40 mA per tube. The tweed Deluxe has 270 Ω, -23 V and more than 40 mA, respectively (depending on rectifier tube and power tubes). With increasing drive level, the average cathode-voltage rises (due to the non-linearity in the tubes), and the operating point shifts towards the "cooler" range. The transfer characteristic becomes flatter. The following measurements were again taken using a stabilized power supply (300 V), with R_{aa} = 6.2 kΩ, no negative feedback, R_{g2} = 470 Ω, ohmic nominal load.

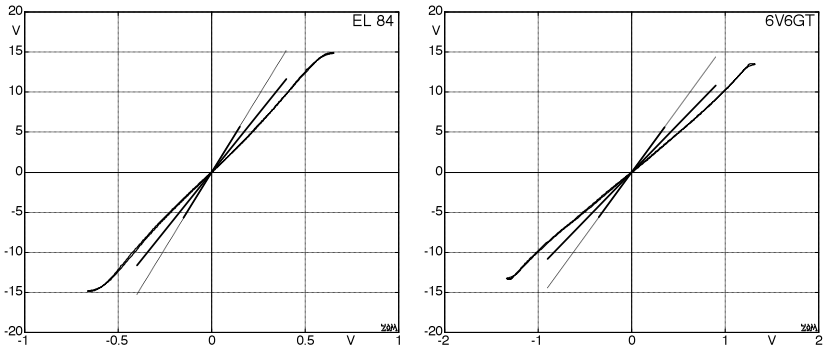

Fig. 10.5.75: Transmission from the phase-inverter to the load-impedance. Power stage with cathode-resistor.
EL84: 120 Ω, 6V6GT: 270 Ω, bridged with 250 µF.
Three different drive-levels.

We can see from **Fig. 10.5.75** that the EL83 and the 6V6GT differ in gain by 7.5 dB, and that the gain drops by 4 dB with increasing drive-level. With a fixed bias-voltage, such a gain-reduction cannot be observed (compare to Fig. 10.5.45). As the overdrive increases, a saddle point in the origin appears for both tubes – here the true pentode differs from the beam-tetrode, though: in the EL84, the characteristic has an almost horizontal slope in the origin (**Fig. 10.5.76**) while for the 6V6Gt, this gain-decrease is much weaker.

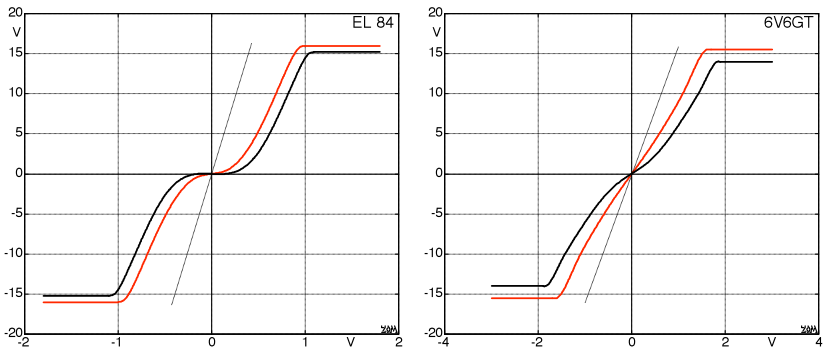

Fig. 10.5.76: Transmission from phase-inverter to load impedance.
With R_K (black) and without (red).
Overdriven power stage.

This *dip* in the transmission curve has several reasons in a guitar amplifier: as the drive-level increases, the supply voltage decreases, and U_{g2} with it; the coupling capacitors towards the phase-inverter change their polarization; if a cathode resistor is present, the voltage drop across it increases. All three effects build up in the same direction and shift the operating point towards the "cooler" range, and consequently the crossover distortion close to the origin increases. Last, the screen-grid resistor needs to be considered, as well: in the EL84, the currents through the screen grid are larger than in the 6V6Gt, and therefore the voltage drop across the screen-grid resistors is bound to be different.

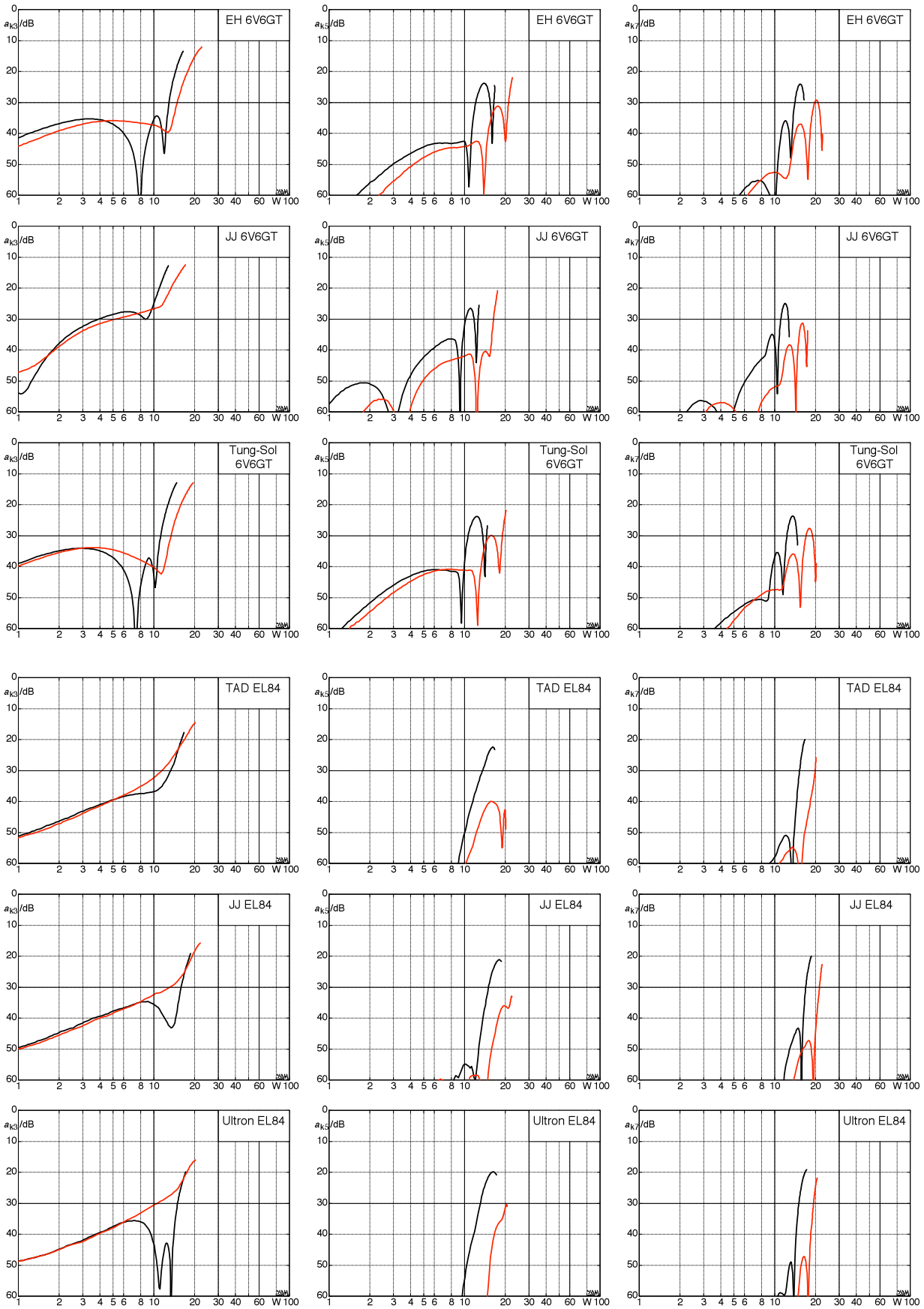

Fig. 10.5.77: Attenuation of distortion a_{k3}, a_{k5} and a_{k7} with nominal load. With (black) and without (red) R_K.

In **Fig. 10.5.77** we see the first three odd-order distortion attenuation characteristics. For k_5 and k_7, the situation is clear: *without* cathode resistor, the power stage distorts more than *with* this resistor, and the EL84 distorts less than the 6V6GT. For the 3rd-order distortion, such tendencies are less pronounced. Strongly overdriven, the amp again has the larger output power without R_K. At around 10 W, however, k_3 is subject to strong fluctuations that are different from tube to tube.

The frequency response of the **internal impedance** is inconspicuous; the measurements did not lay open any significant differences between 6V6GT and EL84. With regard to the operation with and without cathode resistor, no significant differences in the internal impedance could be found, either – as long as the idle-currents were set comparably.

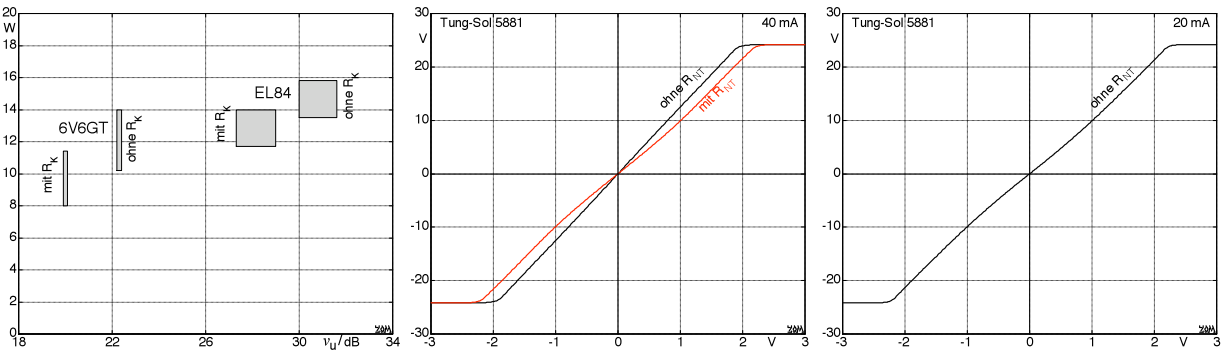

Fig. 10.5.78: Output power vs. gain (left); transmission characteristic (center, right). "ohne"=w/out, "mit"=with.

In the left-hand section of **Fig. 10.5.78**, output power (at 300 V) and gain are shown; again this is only for a small sample. The centre section depicts the effect of the internal impedance of the power supply (R_{NT}). "ohne R_{NT}" (without R_{NT}) indicates the stabilized 400-V-power-supply, being used; "mit R_{NT}" (with R_{NT}) indicates operation from a stabilized 460-V-power-supply, but via a 240–Ω-resistor, and buffered with 47 µF. With the internal impedance of the power supply present, the supply voltage to the power stage drops to 400 V at full power; the bias-current, however, is set for a supply voltage of about 440 V. As the drive-level increases, the operating range wanders off towards "cooler" regions – comparable to the operation without R_{NT}, but with a bias-current of only 20 mA (right). One highly essential difference remains: the red curve is not static! Its slope (= gain) drops with decreasing drive-levels.

Fig 10.5.79 again documents (for a_{k2}) how strong the scatter within new, selected tube pairs can be. Anybody who believes that a THD of 1% is relevant needs to buy better-selected/matched tubes.

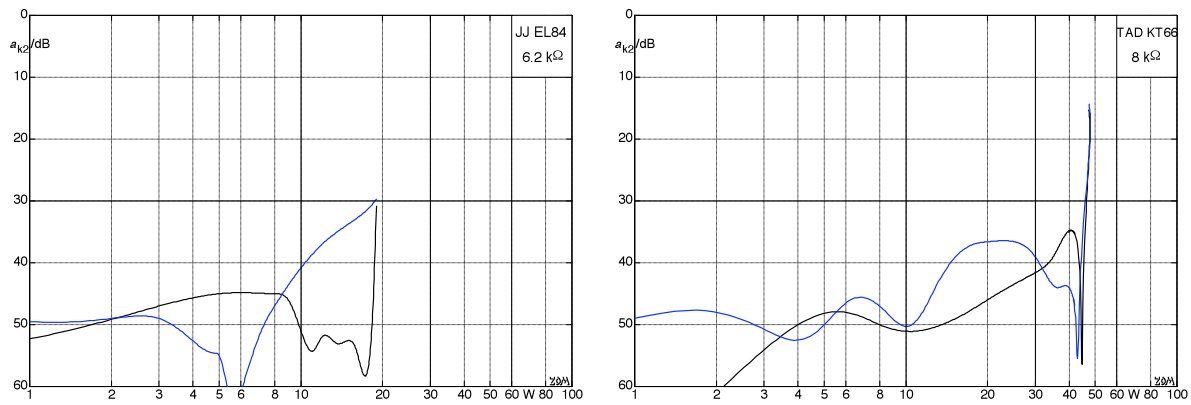

Fig. 10.5.79: 2nd order distortion attenuation; in each picture two newly bought, "matched" tube pairs.

10.5.14 Pentode/Triode/Ultralinear

Pentodes are at work in the power stage of a typical guitar amplifier: EL-34 in the Marshall, EL-84 in the VOX, 6L6-GC in the Fender; or comparable tubes (5881, KT-66, KT-88) – but always pentodes, and not triodes. That some of these tubes actually are beam-tetrodes shall not bother us here, because their beamforming plates are in fact some kind of fifth electrode, as well – even though differences to a true suppressor-grid remain if we apply strict theory. Since these differences are of no significance in the following, we will treat pentodes synonymously with beam-tetrodes.

In a triode, the drive-level-dependent plate-voltage accelerates the electrons, and therefore the gain of the tube is small for low plate-voltage. Conversely, the plate-voltage has only little influence on the emission-current in the pentode because the screen grid is on a high potential independently of the drive-level. The output characteristics of the pentode are therefore almost horizontal (except for the initial distribution area), and the internal impedance is larger compared to triodes. From an overall investigation of efficiency, internal impedance and harmonic distortion, HiFi-developers noticed that both pentode and triode were operating in a sub-optimal border range, and they looked for a compromise. The latter could be found in the **ultra-linear circuit**: here, the screen grid of the power tubes is connected neither to a constant potential (pentode operation) nor to plate potential (approximately corresponding to triode operation), but in between. Since all voltages between supply voltage and plate-voltage are available at the output transformer, it is merely necessary to include a suitable "tap" from the primary winding. This is why an ultra-linear output transformer does not have three but five connections on its primary side. The back-channeling of the signal to the screen grid (g_2) has the effect of a negative feedback that was seen as advantage in HiFi-amplifiers. It appears the same did not happen for guitar amps since only few experiments made the jump to production, such as the 1979 Twin Reverb, some Sunn-variants or – will wonders never cease! – the **200-Watt-Marshall**: yep, there's an ultra-linear power stage, designed for the least amount of distortion. Later, though, the JCM-800-series amps went after their business again without ultra-linearization.

Due to the reduction of the screen-grid-voltage, the obtainable maximum output power drops. This can be used to convert a 100-W-amp into a 50-W-amp. Two switches are installed at the screen grids such that either the full supply voltage or the plate-voltage is connected to the screen grids. Sure, it would also be possible to simply reduce the gain if it gets too loud, but power-stage distortion happens only when overdrive occurs. If the screen grids are connected to the (corresponding) plates, the power pentodes operate in a kind of **triode-mode**: with smaller maximum power, but also with smaller internal impedance. The switch from pentode- to triode-mode therefore does not only change the maximum power (loudness) but also the sound. The operation with high internal impedance emphasizes treble and speaker-resonances, and in the triode-mode the sound looses brilliance and volume. To which extent this is in fact audible depends on (besides the screen-grid-voltage) the negative feedback of the power stage. Changing the screen-grid-voltage implies changing the gain (i.e. the loop-gain) and therefore changing the negative-feedback-factor. It may consequently be that, besides the screen-grid-voltage, the NFB-loop needs to be switched as well. The pentode/triode-switch has no bearing on the **operating point** because at idle, the plate-voltage is almost the same as the supply voltage (the primary winding is of low-impedance of DC current).

Fig. 10.5.80 shows an example for the difference between pentode and triode. An **EL34-power-stage** with Marshall-transformer (JTM-50) is operated with a stabilized voltage of 400 V, and a Marshall-Box 1960-AX as loudspeaker. First, the investigation targets the influence of the screen-grid resistor at 1.5 kΩ, 470 Ω, or 0 Ω. A small effect shows up with the gain: for R_{g2} = 0 Ω, the gain is 1.3 dB larger than for R_{g2} = 1.5 kΩ. The strain on the screen grid is heavily affected: with overdrive, the screen grid **glows** barely visibly♣ for R_{g2} = 1.5 kΩ, but lights up to **bright red** with R_{g2} = 470 Ω, and to **bright yellow** at R_{g2} = 0 Ω**!** In the interest of a long tube life, a sufficiently large R_{g2} should always be used. The flipside is that the power stage generates, with R_{g2} = 0 Ω, 1/3 more power compared to R_{g2} = 1.5 kΩ. Let's now look at the figure: the triode configuration reduces the internal impedance, which makes the gain drop in a particularly strong manner for a high-impedance load (Chapter 11.2). The gain is smaller by about 5 dB at 400 Hz in the triode-mode; in the figure, this was balanced out for small drive level (normalization to 400 Hz).

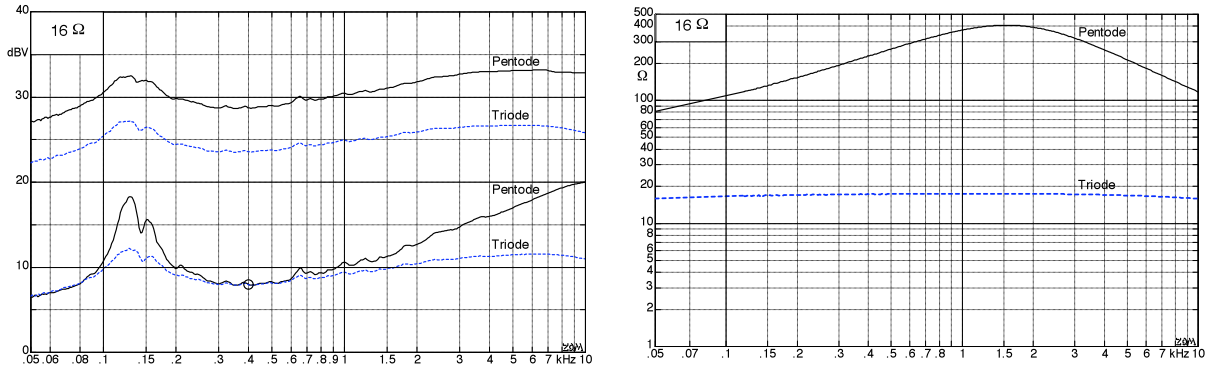

Fig. 10.5.80: Voltage transmission for pentode- and triode-operation of the power tubes (EL34, 1960AX).
The measurement curves at small drive level were normalized to 400 Hz; no normalization was done for high drive level. On the right the frequency responses of source-impedances of the power stage are shown (internal impedance at the output transformer). For all these measurements, the negative feedback was deactivated.

At small drive level (i.e. for linear operation), the power tubes are of high impedance (about 30 kΩ) in pentode-mode, and the source impedance measured at the output is co-determined by the output transformer. In triode-mode, the internal impedance of the tubes drops to about 1.2 kΩ: now, the source-impedance is predominantly determined by the internal impedance of the power tubes. As the power stage is driven to the extent that limiting occurs (measurement curves for high levels), we see the differences in maximum power output. On the other hand, the characteristics of the curves clearly become more similar. **Conclusion:** In triode-mode, brilliance and emphasis of the speaker-resonance drop for undistorted operation. The maximum power-output drops to about 1/3, and when overdrive occurs, the frequency responses become similar.

For the **power stage with active negative feedback**, the differences in the frequency responses is much smaller for linear operation while the differences in the maximum power output remain similar. It is well known that negative feedback has no effect on the maximum power yield: the non-linear distortion is reduced somewhat but the limit values of the tubes cannot by modified via NFB. However, even merely moderate NFB decreases the source impedance so strongly that differences in frequency response between triode- and pentode-mode become meaningless. **Simply put**: negative feedback transforms the power stage from a current source to a voltage source.

♣ As seen with the JJ-EL34; different strain-situations can occur with other tubes.

Fig. 10.5.81 shows how gain and source impedance are reduced by the negative feedback. The **NFB-factor** (= 1 + loop-gain) depends, for a tube amplifier, on the load (speaker impedance). Measurement and calculation (circles at 400 Hz) match very well. For an NFB-factor of 1, the power stage has no negative feedback, while an NFB-factor of 6.73 already represents a strong feedback for power stage based on tubes. The frequency characteristics of **pentode- and triode-mode** become very similar already for moderately strong NFB, and therefore in practice all that remains is a small difference in gain.

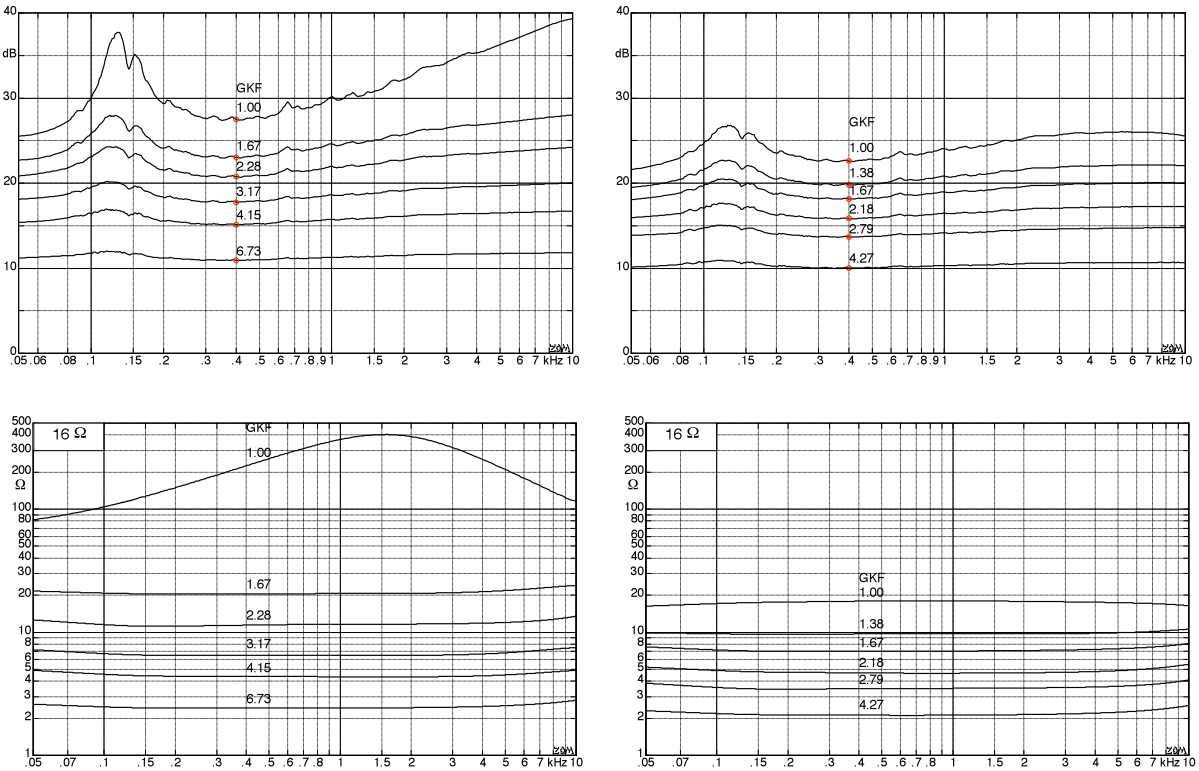

Fig. 10.5.81: Gain from the input of the phase-inverter to the output of the transformer (16 Ω, 1960AX).
Left: pentode-mode, right triode-mode. Lower line of pictures: source impedance; "GKF" = NFB-factor.

Besides the drop in output power and the change in frequency response, switching from pentode- to triode-mode brings a further consequence: the non-linear behavior changes. The triode-characteristic has multiple bends while the pentode characteristic is degressive. This has an effect on the **distortion-attenuation**: a reversal of the sign of the curvature (2nd derivative) leads to a zero in the harmonic distortion in triode-mode (**Fig. 10.5.82**).

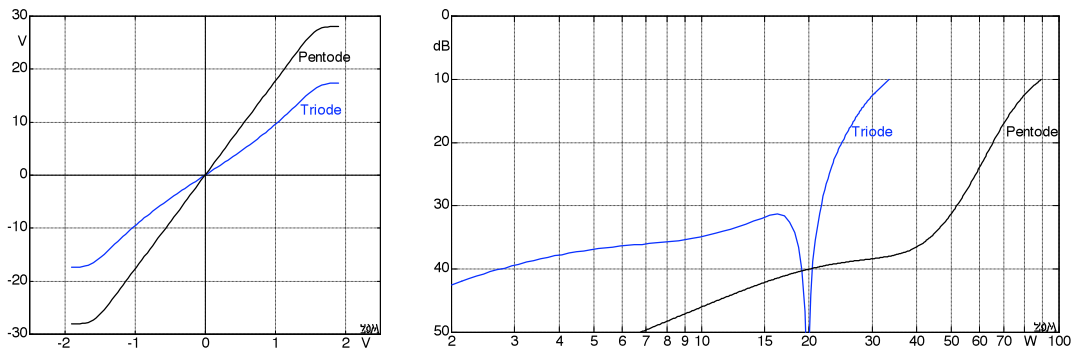

Fig. 10.5.82: Transmission characteristic (left), 3rd-order distortion attenuation (right). 2xEL34, nominal load (ohmic), no NFB. The exact curve depends on the individual tubes and on the bias-current.

10.5.15 … and the current flows on while you are long dead

Suggestions regarding modifying a tube power-stage will entice to do just that. Swap your power tubes, install different filter caps, modify that negative feedback. To cite myself: The fact that not everybody who removes an amp chassis from a cabinet instantly keels over dead must not lead to the conclusion that this will never happen [Chapter. 10.5.8]. A tube amp operates on the basis of life-endangering voltages, any musician screwing (sic!) with it, as well. Therefore, let me repeat: working on a tube amp requires a specialist education. And even if the courageous/experienced/lucky customizer is left unharmed: it is bad enough if the power transformer gets fried. Or if the loudspeaker expires right in the midst of the most important solo of ones life … because it could withstand the original 40 W, but decided to succumb to those after-mod-80 W.

Books, magazines, and fora on the world-wide-web, are filled with recommendations how to customize your amp. More crunch, more bass, more treble, more oomph, more of everything. Swapping the output transformer can lead to additional strain on the mains transformer and it can overload the rectifier tube (if one is in the game). The expert can size up all this but the layperson can't. 6L6-GC and KT-66 may be swapped for each other, as long as the bias-current is correctly adjusted afterwards. A change from the 6L6-GC to the EL34 represents already a potential power increase, and needs to be carefully considered and implemented. The socked-connections need to be checked when doing this because there are differences here. Power stages with the 6V6 are particularly dangerous candidates: whoever – hoping for the triple output power – plugs in two 6L6-GC (or even EL34's) instead of the 6V6-GT acts negligently. If the object of experimentation were a tweed Deluxe, we would first need to have a look at the rectifier tube: the 5Y3-GT is a good partner for the 6V6-GT, but not for the EL34. So that needs to be changed, too: instead of the 5Y3-GT, the 5U4G gets to be plugged in – or should the GZ34 be used, yielding limitless current? But then there's the cathode resistor: 270 Ω. The EL34 easily exceeds a cathode-current of 300 mA, so that's 24 W dissipated in the cathode resistor. Which actually is a power increase of some kind – but probably not the desired one. For the Deluxe Reverb, this problem disappears: there is no cathode resistor. Still, the mains transformer needs to be watched: it not only needs to supply an additional 2 A of filament-heating current, but also the desired additional output power. Likely to be forgotten is the increased power dissipation in the tubes: it's about 15 W for the 6V6-GT but double that for the EL34. The EL34 is a true pentode but the 6V6-GT is not.

Those who are "in the know" can do such conversions. But then you read in a forum: *my new transformer has a wire more than the old one … what should I do?* Or: *the big resistor is shot. Where can I get a new one? Or can I just leave it out*♣*?* Simple answer: HANDS OFF!! You don't get a kidney transplant done in your auto-shop, either, now do you?

♣ In fact, what we read is: *That biggie resister is in ashees where do i get anew 1. Or cann I jus leave it of?*

