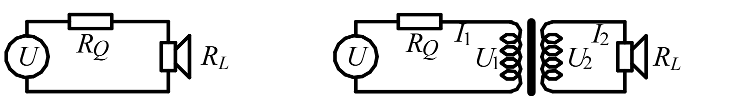
10.6 Output transformer

Typically, your customary power tube will have an optimum load-impedance in the kilo-ohm-range i.e. about 1000 times the impedance of a loudspeaker. If, for example, an 8-Ω-load-impedance were to be connected to a source having an internal impedance of 8000 Ω, then 99,9% of the generated power would be dissipated via the internal impedance, and only 0,1% would arrive at the load-impedance. That is of course not acceptable. Tubes operate at high voltages (400 V) but can digest only small currents (0.2 A). With loudspeakers, the situation is exactly the other way round: a 4-Ω-loudspeaker requires 16 V to take on 64 W, with a current of 4 A flowing through it. The output transformer (OT) has the task to **match** the high-impedance tube circuit to the low-impedance loudspeaker. As a matter of principle, the OT at the same time works as a filter that rejects high and low frequencies, and it generates special non-linear distortion. While the matching function of the OT is relatively easily calculated, the non-linear distortion eludes an exact description. The corresponding models are therefore either inadequate, or not at all readily understood, or both. The following elaborations try to give a clear picture on the basis of specific measurements. For the latter, genre-typical output transformers were used – they do, however, not represent any selected sample-median.

10.6.1 The linear model

Impedances (complex resistances) are only defined within the linear model [20], and therefore the impedance transformation can be calculated only for a linear output transformer. The AC-source is the tube circuit that is assumed to be a voltage-source with a (series-connected) source-impedance R_Q. The load is given by the loudspeaker-impedance R_L (**Fig. 10.6.1**), and both source- and load-impedance taken to be purely ohmic for our first investigations.



$$U_1/U_2 = TR = I_2/I_1$$
$$U_1 \cdot I_1 = P = U_2 \cdot I_2$$

Fig. 10.6.1: AC voltage-source with load-impedance; with & without an ideal matching transformer.

The transformer shown here is of **ideal** characteristics, and completely described by the two equations given above; $TR = N_1/N_2$ is the **turns-ratio**, also termed **transformer-ratio**. The windings shown in the schematic therefore must not be interpreted as inductances but have a purely symbolic character. The **idealization** mentioned above may be in sharp contrast to reality: the ideal transformer can transmit DC – something impossible for a real transformer. For our first forays into transformer-land, this discrepancy is not a problem – we can (and will have to) expand the **model** as needed. According to the idealization, the transformer is also loss-less: $U_1 \cdot I_1 = U_2 \cdot I_2$. In the interior, energy is not stored, nor dissipated into heat. This is another difference to the real transformer: its windings do generate heat – which is not (yet) considered in this simple model. The latter is not able to simulate the non-linearity (magnetic hysteresis) caused by the iron core, and the same holds for winding capacitances and leakage flux. All these specific characteristics will need to be incorporated in a realistic model, and we can already now anticipate how complex this is likely to become.

The power matching, on the other hand, may very well be shown using the ideal transformer: the source (voltage-source with source impedance) "sees" as load the input-impedance R_E of the output transformer (**OT**):

$$R_E = \frac{U_1}{I_1} = \frac{U_2 \cdot TR}{I_2 / TR} = \frac{U_2}{I_2} \cdot TR^2 = R_L \cdot TR^2$$ Impedance-transformation

The secondary load-impedance (R_L) is mapped (transformed) via the OT into the primary input-impedance of the OT. If this input-impedance R_E is very small relative to R_Q, the major part of the power is fed to R_Q, and not to R_L. Conversely, if R_Q is large, almost all power is fed to R_L, but due to $P \sim 1/ R_E$, this power becomes smaller as R_Q becomes larger. Therefore, equal internal- and load-impedance is often sought as an **optimum for matching**: $R_Q = R_E$. With internal impedance and load-impedance known, the transformer-ratio can easily be calculated from this simple condition: $TR = \sqrt{R_Q / R_L}$. Given R_Q = 7200 Ω and R_L = 8 Ω, we get, for example, a transformer-ratio (turns-ratio) of TR= 30 (tube amplifiers Chapter. 10.6.2).

So, how exactly does the output transformer accomplish this transformation, how does it generate the secondary quantities from the primary ones? This is done via the magnetic coupling of two windings the turns-ratio of which corresponds to the transformer-ratio TR. The primary current I_1 flowing through the primary coil generates a **magnetic field** that, in an ideal transformer, entirely permeates the secondary winding and induces the secondary voltage U_2. If the transformer has a load coupled to its secondary winding (as it is normally the case), there is also a current in the secondary circuit that itself generates a magnetic field entirely permeating the primary winding (in the ideal transformer) and inducing a voltage there. Both coupled processes (current → field → voltage) can and need to be superimposed; this is the basis for the calculation of the general case [4, 7, 17, 18, 20]. However, a wire configured as a winding needs to be represented in the **equivalent circuit diagram (ECD)** at least via a resistor (copper-resistance) and an inductance (magnetic field) – which leads to a first extension of the ideal transformer-schematic. Since the magnetic coupling of the two windings is an indispensable basis, it needs to find its way into the transformer-ECD, too. How this ECD is derived from the physical interrelations shall not be elaborated here explicitly – extensive literature already exists for this (see above). Basically, the real transformer can be represented by a special ideal transformer and several supplementary two-poles. The special ideal transformer is fully described by its transformation ratio TR, and what has been stated in Fig. 10.6.1 does hold for it. The supplemental two-poles approximately model the characteristics in which the real transformer differs from the ideal one. Still: these are approximations the applicability of which needs to be checked in each individual case.

The most important characteristics modeled by the supplemental two-poles are: resistive losses, inductances, and flux-leakage. Losses are due to the copper wire and the magnetic core, inductances result from (coupled) windings, and flux-leakage happens because, in the real transformer, not the whole magnetic flux generated by one winding permeates the second winding, but a part misses it. The **leakage-factor** σ defines the extent of the flux-leakages; alternatively, the **coupling-factor** $k = \sqrt{1-\sigma}$ can be given. A leakage-factor of σ = 0% corresponds to complete coupling (= ideal tight coupling), while a leakage-factor of 100% indicates non-coupled windings. There are different equivalent circuit diagrams; the individual factor TR may deviate from the physical turns-ratio.

Two of the most important ECD's are shown in **Fig. 10.6.2**. R_1 and R_2 represent the ohmic components of the winding-impedances and model the copper-resistances. L_1 and L_2 are the inductances of the primary and the secondary windings, respectively. For a secondary open-circuit, the measurement of the primary input-impedance yields $R_1 + j\omega L_1$. For a primary open-circuit, the measurement of the secondary output impedance yields $R_2 + j\omega L_2$. The inductance designated M in the right-hand ECD is the **mutual inductance**. The following relationships hold: $M = k\sqrt{L_1 L_2}$, $k = \sqrt{1-\sigma}$, $TR = N_1/N_2 = \sqrt{L_1/L_2}$.

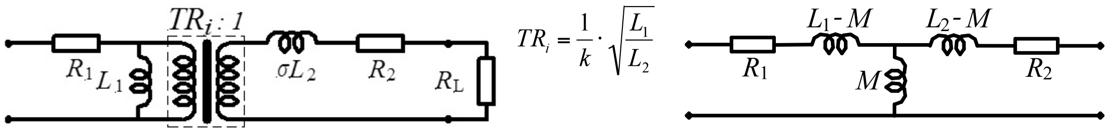


Fig. 10.6.2: ECD's for transformers. The transformer in the ECD on the left is ideal (and thus free of inductances). The inductances in the ECD on the right may become negative; this does not restrict the validity.

Besides the three ohmic resistances that can be easily determined from a DC-measurement, the ECD holds **three degrees of freedom**: L_1, L_2, and k. L_1 and L_2 may be ascertained e.g. via an impedance-measurement with contra-lateral open circuit. The coupling-factor can be determined with contra-lateral short-circuit. Measuring the primary DC-resistance R_1 of the OT is most unproblematic, while regarding the secondary resistance we need to bear in mind that it may by of very small magnitude (possibly $R_2 < 0.1\Omega$). When measuring the inductance, the fact that the ECD mentioned above has only limited applicability in practice requires consideration: stray- and winding-capacitances influence the impedance, as well (Fig. 10.6.4).

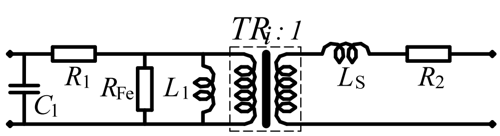
In both ECD's given in Fig. 10.6.2, the inductance in the parallel branch will short any DC voltages – the result is a **high-pass**. Accordingly, the parallel inductance needs to be as large as possible in order to allow for low-frequency operation. The inductance rises approximately with the square of the turns-number of the winding, and therefore a winding with a high turns-number would be desirable – however, this brings along mounting copper-resistance, and correspondingly increasing losses. To keep the copper-resistance low, the cross-section of the deployed wire needs to be large – requiring the dimensions of the transformer to be large, as well. **Simple conclusion: transformers that handle high power and low frequencies need to be large**. For the selection of the cross-section of the wire, the **current-density** supplies a first step of orientation: given an RMS primary current of 0.11 A, a 0.2-mm-wire would be suitable for 3.5 A/mm^2. The latter value is just for orientation: for large transformers, somewhat smaller current-densities will have to be assumed, especially if the surrounding air is heated up by the tubes. The current I_2 flowing in the secondary winding is larger than the primary current I_1 by the factor of TR; however, the secondary turns-number is $1/TR$-fold smaller than the primary turns-number; the product of current-strength and turns-number therefore is the same for primary and secondary winding. This holds at least for the ideal transformer – in real transformers there are small deviations that may, however, be disregarded for a first consideration. Given equal current-densities for primary and secondary winding, it follows from the equation $I_1 N_1 = I_2 N_2$ that the **cross-sectional areas of the windings** should be equal for both windings. The total cross-sectional area of the winding (amounting to e.g. 2.2 cm^2 for the M55-transformer) therefore is made available with 50% each to both primary and secondary winding. Depending on the application, transformers need to meet certain requirements, for example with a proof-voltage of more than 1000 V (and corresponding supplementary insulation layers), or a special low-capacitance winding (with different build), or additional taps (requiring more contact wires and thus space). This shows that transformers may have manufacturer-specific differences that are not obvious at first glance.

The M55-transformer cited as an example has a winding-surface of 2.2 cm^2 i.e. 1.1 cm^2 per winding. This value must, however, not be simply divided by the cross-sectional area of the wire because wire-insulation and -spacing also require space. Nevertheless, it should just about be possible to accommodate 2000 turns of 0.2-mm-wire. Applying the current (e.g. 0.11 A) as calculated from the current-density yields a magnetomotive force of 220 A, and a magnetic field-strength of 1.7 kA/m (as a first-order approximation). From a thermal point-of-view, this may be o.k. – from a communication engineering point-of-view, it is not: the materials normally used for cores in transformers are all but "saturated" at such high field-strengths, and the magnetic flux cannot increase anymore if the field-strength is further increased. Strong non-linear distortion would be the result. Schröder recommends in Vol. 1 of his book *Elektrische Nachrichtentechnik* a maximum magnetic field-strength of 0.1 kA/m. Consequently the overdrive found in our above example would be massive. Alternatively, the **maximum magnetic flux-density** could also be calculated:

$$\hat{B} = \frac{\sqrt{2} \cdot \tilde{U}_1}{2\pi f \cdot N_1 \cdot A_{Fe}}$$

Peak value of the magnetic flux-density.
N_1 = primary turns-number,
A_{Fe} = cross-sectional area of iron.

It is clear from the reciprocal dependency on frequency that, for a primary voltage U_1 sourced form a stiff voltage-source, the flux-density decreases with increasing frequency – therefore problems may result in particular for low frequencies. We will get back to the non-linear behavior in Chapter 10.6.4; first, the behavior for small drive-levels is under scrutiny. The (linear) ECD's introduced in Fig. 10.6.2 enable us to approximately describe impedances and transmission behavior of an output transformer. In the higher-frequency region, however, noticeably deficits remain because capacitive coupling among the windings and iron losses are not considered yet. Strictly speaking, every differential section of the winding is capacitively coupled to every other section, but a *single* **substitute capacity** is sufficient to model this infinite number of coupling capacitances. The **iron losses** (hysteresis- and eddy-current-losses) may be modeled via a resistor with good approximation, as well, and an extended equivalent circuit diagram shown in **Fig. 10.6.3** represents a good compromise between complexity and accuracy. Calculations with the approximation $TR_i \approx TR$ are always acceptable: the transformers considered here rarely have a leakage-factor of in excess of 1%.



C_1 = capacitance of the winding,
L_1 = primary inductance,
R_1, R_2 = copper-resistances,
R_{Fe} = iron losses,
L_S = leakage inductance.

$$L_S = \sigma \cdot L_1 / TR^2$$

$$TR_i = TR / \sqrt{1-\sigma}$$

Fig. 10.6.3: Equivalent circuit diagram of transformer♣ (linear model). Non-linear behavior: see Chapter. 10.6.4.

Fig. 10.6.4 shows comparisons between measurements and calculations carried out on the basis of the above model. Since all these transformers are used in push-pull output stages, the respective primary winding is divided in two halves. Calculation and measurement was respectively done for one half of the primary winding. For secondary open-loop operation, the primary impedances of the two winding-halves are practically identical; there are differences for secondary short-circuit, though – these are due to different coupling of the windings. For low-impedance loading (i.e. for loudspeaker-loading, as well) the push-pull drive-signal therefore is not symmetrical anymore in the higher-frequency region.

♣ The capacitance may also by connected on parallel to L_1; the differences are small.

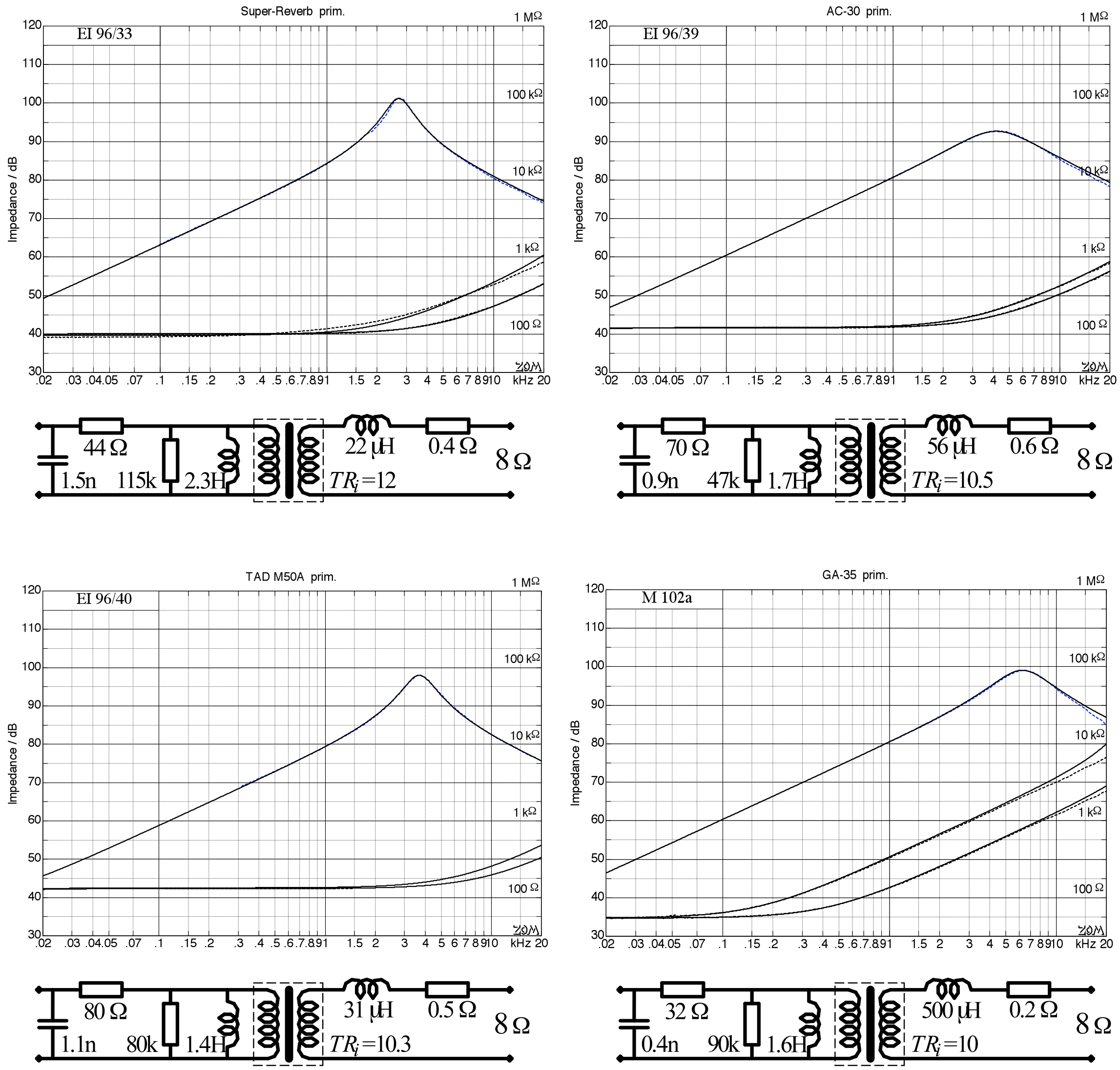


Fig. 10.6.4: Comparison of impedance measurements (------) and model calculations (———), each for one half of the primary winding (R_a). The two open-loop impedances are practically identical; the short-circuit impedances differ due to different coupling-factors.

Measurements and calculations in Fig. 10.6.4 are practically identical over a wide range but there are some sections in which differences become apparent. In principle it would not be difficult to extend the model by a few further components such that a good correspondence would be achieved across the whole frequency range. However, in the interest of general applicability, the ECD as developed above shall remain unchanged. The divergences are rather limited, anyway.

We can also see from Fig. 10.6.4 that – at least for the transformers investigated here – the ECD is well suited to model the primary load-impedance (i.e. the strain on the power tubes) **for linear operation**. However, output transformers work linearly only for *very* small output power, typically $P < 1$ mW. For your regular output power, the **parallel inductance** (L_1), in particular, depends very strongly on the drive-level. As simple as the linear equivalent circuit diagrams are, their applicability still remains strongly limited. For this reason, Chapter 10.6.4 will elaborate more extensively on the non-linear behavior.

10.6.2 Impedance-matching and transmission

Frequently, the term "impedance matching" is interpreted such that, for a maximum of power-yield, the source- and the load-impedances need to be equal (or conjugate). The datasheet of the power-tetrode 6L6-GC lists an internal impedance of 35 kΩ so that we could conclude that the primary impedance of the output transformer should also amount to 35 kΩ. At the same time, however, the datasheet specifies a so-called "optimum load impedance" at no more than 1.4 kΩ. What follows is this: the 6L6-GC is (like all tetrodes♣) a high impedance source and operates approximately as a current source. The power delivered by a current source is proportional to the load-impedance: the higher the latter the higher the power-yield. However, this simple relation is limited by three non-linear conditions: the maximum allowable plate-dissipation, the maximum allowable plate-voltage, and the residual voltage at the plate. The **optimum load-impedance** (= external impedance) results from these non-linear conditions, and not from the equality of internal- and load-impedance. It is sufficient, as a rule, to assume the internal impedance of the tube to be large relative to the load-impedance; the optimum load-impedance (per plate) for push-pull stages usually is about 1 – 2 kΩ.

The output transformer enlarges the secondary load-impedance (typically, this is the loudspeaker impedance) by the square of the turns-ratio, for example:

An 8-Ω-load-impedance is transformed – for $TR = 12$ – into 144 x 8 Ω = 1152 Ω.

Usually, there is no need to distinguish between the turns-ratio of the windings $TR = N_1/N_2$, and the transmission ratio TR_i in the equivalent circuit diagram, because in most cases the respective values differ by less than 1% (Fig. 10.6.3). The internal impedance R_i of the tube is transformed with TR^2, as well: the internal impedance of the replacement source driving the loudspeaker amounts to R_i / TR^2 (in the example 35 kΩ / 144 = 243 Ω). As long as the power stage is not overdriven, it will operate the loudspeaker approximately as a **stiff current-source** – if the power stage does not involve **negative feedback** (NFB). The voltage/voltage-NFB implemented in many amplifiers reduces the internal impedance of the power amplifier. Still, perfect behavior as a stiff voltage-source is not accomplished by tube power-amps (however, most transistor power-amplifiers will achieve this – but they are not a object of the present investigations).

Fig. 10.6.5 shows the family of output characteristics for a power-pentode known from Chapter 10.5, plus some load-dependent transmission characteristics. Given the secondary impedance (e.g. 8 Ω), the slope of the operating characteristic may be changed as needed.

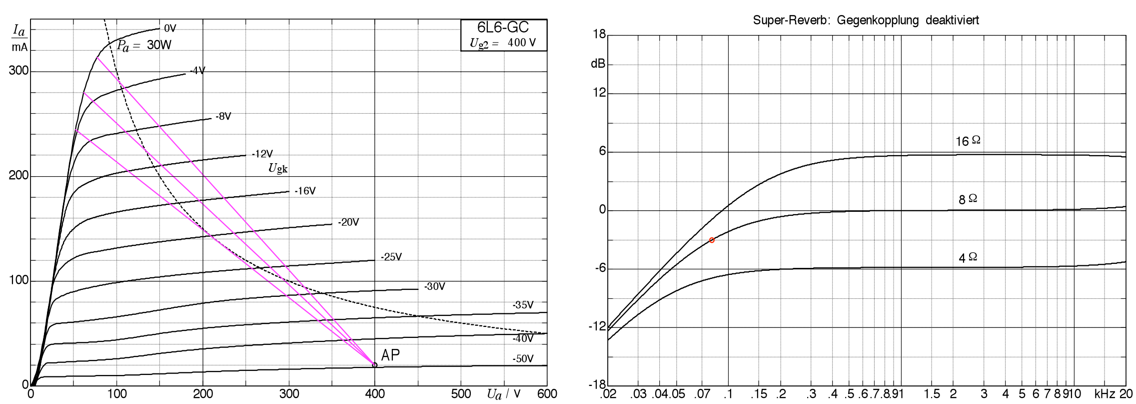


Abb. 10.6.5: Transmission characteristics (left), frequency-response at the 8-Ω-output for a load of 4/8/16 Ω.

♣ As far as they are not operated in triode-mode (triode-mode: g2 and plate are directly connected).

It may be matched to the family of characteristics discretionarily by varying the transmission ratio (*TR*): a larger *TR* results in a flatter curve for the load-line i.e. a smaller plate-current and a larger voltage swing.

The internal impedance of the tube transformed via TR^2 is, however, not the source impedance relevant for the loudspeaker across the whole frequency range. The equivalent circuit diagram presented in Fig. 10.6.2 shows that the parallel inductance L_1 determines the impedance at low frequencies: it shorts the source for low frequencies and has the effect of a **high-pass**. Moreover, we need to consider that this inductance is **non-linear**, and therefore we do not have a conventional high-pass here (Chapter 10.6.4). The transmission curves given in Fig. 10.6.4 involve a demagnetized transformer core; however, this can be achieved only at untypically small drive-levels of **about 1 µW**. Nobody will play a 45-W-amp at such a small power level – the tube amp will not be able to shape the sound in the way for which it is designed. Still, the curves shown in Fig. 10.6.5 had to be measured approximately at this power level, otherwise the main inductance L_1 would have become dependent on drive-level in a rather unbecoming way. The small-signal ECD so popular in communication engineering it in a bit of trouble due to this, but it can be rescued by a special modeling at low frequencies (Chapter 10.6.4). Basically, the parallel inductance looses its impact with rising frequency, and the transmission becomes frequency-independent (for an ohmic load). At very high frequencies (that can however barely, if at all, be reproduced by a typical guitar-loudspeaker), the incomplete field-coupling and the winding-capacitances may start to have an effect – but in all likelihood this will not be dramatic or noticeable at all.

Power amplifiers are always specified for a real (ohmic) **nominal load-impedance** although the impedance of a loudspeaker is always dependent on frequency. For this reason, **Fig. 10.6.6** depicts transmission frequency responses for loading with a loudspeaker; the mapping of the frequency-dependent loudspeaker impedance onto the frequency response is clearly visible. The power stage of a Super-Reverb normally has negative feedback but for these measurements it was deactivated – otherwise the characteristics of the output transformer would have been suppressed too much (operation with negative feedback: Chapter 10.5). The operation with a loudspeaker results in a treble boost (voice-coil inductance), and between 10 and 100 Hz we observe a narrow-band boost due to the loudspeaker resonance. For both operational states, attenuation shows up in the bass range for very small drive-levels ($P <$ 1mW): this is due to the main inductance (see also Chapter 10.6.4).

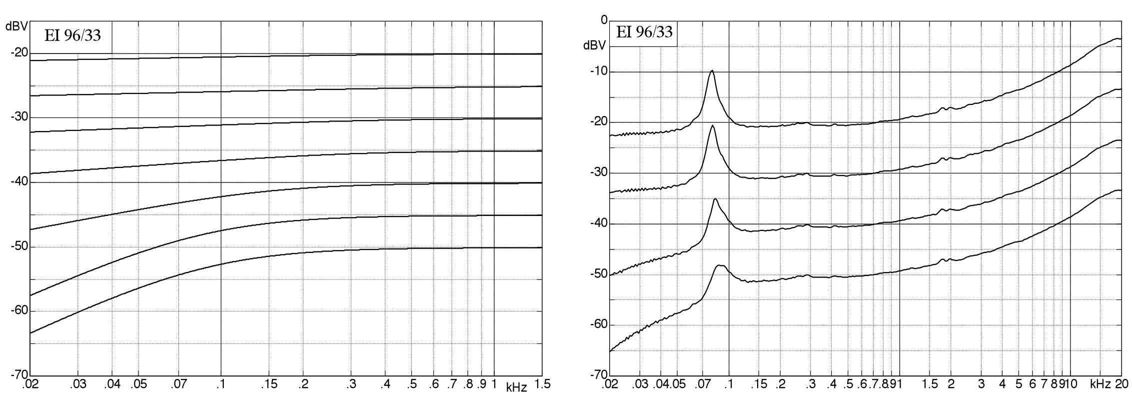


Fig. 10.6.6: Transmission frequency response; transformer with a secondary load of 8 Ω (left), and loaded with a real loudspeaker (right). NFB deactivated. 8-Ω-load yields a voltage level of –20 dBV $\Rightarrow$ P = 1.25 mW.

10.6.3 Winding-capacitances & -asymmetries

In order for the push-pull power-stage to assemble the two half-waves of the signal correctly with respect to magnitude and phase, the primary windings of the transformer need to be completely similar. Which of course they are not, because they cannot be located at one and the same position on the winding-former. If first one primary winding is wound, and then the second on top of the first, the difference in wire-length is immediately apparent. Furthermore, measurements in the high-frequency range will reveal differences in the coupling- and leakage-factors, and in the winding-capacitance. To moderate these problems, the windings are subdivided (**Fig. 10.6.7**), and the subsections are alternately wound on top of each other (or next to each other in **multi-chambered** transformers).

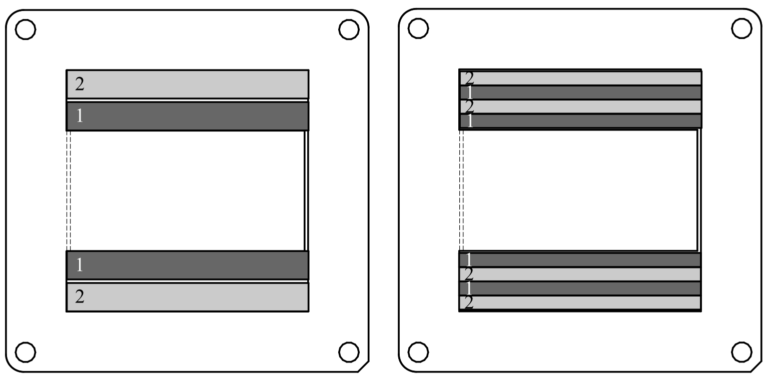


Fig. 10.6.7: Construction of the winding. In the interleaved winding (right), the sub-sections of different windings alternate. In transformers with a sophisticated build, we find multiple "nestings" of primary and secondary winding.

In the *RL*-equivalent-circuit-diagram of the transformer (Fig. 10.6.2), the relative **bandwidth** (f_H / f_T) is inverse to the leakage-**factor**; with a favorable build of the winding three frequency-decades can be covered which is sufficient even for HiFi-quality. However, the winding capacitance must not be completely ignored – in order to describe the high-frequency transmission characteristic, at least *one* capacitance is required (e.g. Fig. 10.6.3). It is this capacitance that determines (together with other parameters) the upper cutoff frequency, and it is just as important as the stray-inductance. As an example, two transformers were examined that are both offered for the **Fender Tweed Deluxe**: the 1750E from Hammond and the TAD-1839. **Fig. 10.6.8** shows the transmission frequency responses measured for loads of 8 Ω and 80 Ω at the secondary output (with a stiff current source driving *one* primary winding). Both transformers show a resonance-emphasis at high frequency: the effect of stray-inductance and winding-capacitance. Since loudspeakers do not merely represent simple ohmic resistances (Chapter 11), supplementary measurements were taken with an 80-Ω-load. This suddenly revealed serious differences, and consequently specifications at nominal load are a necessary but insufficient criterion.

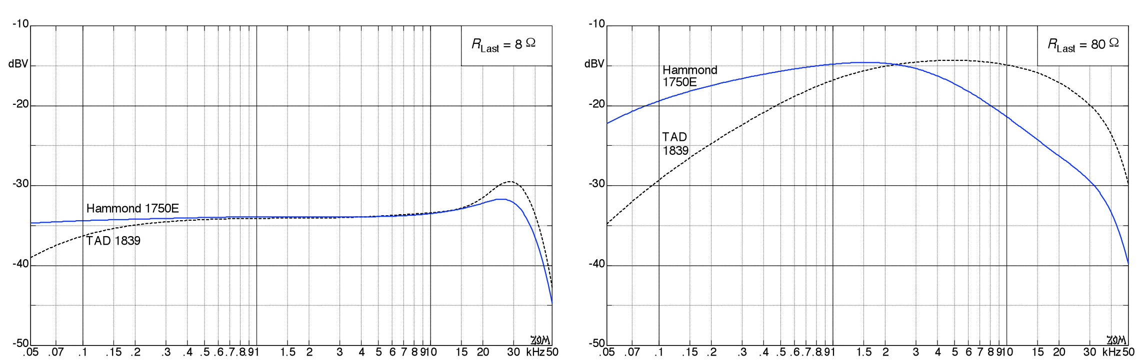


Fig. 10.6.8: Frequency response with a stiff current source (0.16 mA) driving one primary winding.

A short diagnosis of Fig. 10.6.8 could read: *the Hammond lacks in treble, and the TAD lacks in bass*. That is too simplified, though, and we need to dive a bit more into the details. The measurements in fact happen at a rather small primary current and, according to Fig. 10.6.6, the main inductance (see Fig. 10.6.6) is relatively small here. Also, a loudspeaker impedance of 80 Ω is, in reality, not actually reached at high frequencies. Therefore, supplementary measurements are required with loading by a real loudspeaker. These are shown in **Fig. 10.6.9**, with a **Jensen P12N** (mounted in a Deluxe-cabinet) loading the output transformer. Using a stiff current-source again reveals a slight deficiency of the TAD-transformer in the bass-region although this becomes less significant as the drive-level increases. The treble-deficiency of the Hammond-transformer remains relegated to ranges which – for a 12”-speaker transmitting frequencies up to about 5 kHz – have no practical bearing. Our revised conclusion therefore is: in the transmission range important for electric guitars, the Hammond 1750E offers a marginal advantage versus the TAD-1839 – this would possibly justify a small mark-up for the Hammond. Surprise, though: at the time of this writing (AD 2012), TAD charges a stout 86,20 Euro for the 1839 while the Hammond 1750E sets you back a mere 34,70 Euro at Tube-Town. Both TAD and Tube-Town offer a whole range of further output transformers; Chapter 10.6.5 includes corresponding measurement results.

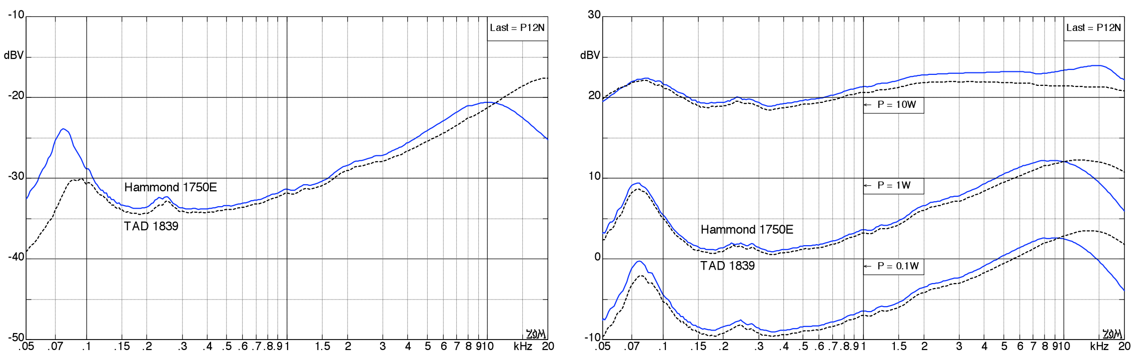


Fig. 10.6.9: Frequency responses with loudspeaker-loading: stiff current-source (left), power-stage (right). 20 dBV at 8 Ω yield => P = 12.5W, P = 10 W corresponds to a voltage level of 19 dBV. At voltage levels around 20 dB, this 6V6-GT-power-stage already shows significant non-linear distortion.

Figs. 10.6.8-9 show the transmission from *one* primary winding to the secondary winding – there are, however, *two* primary windings that feature different magnetic and capacitive coupling to the secondary side. **Fig. 10.6.10** considers this and shows both transmission functions. Again, it becomes apparent that an ECD of pure *RL*-build is not adequate, although the figure also clarifies that the differences are limited to ranges that are not relevant for guitar amplifiers.

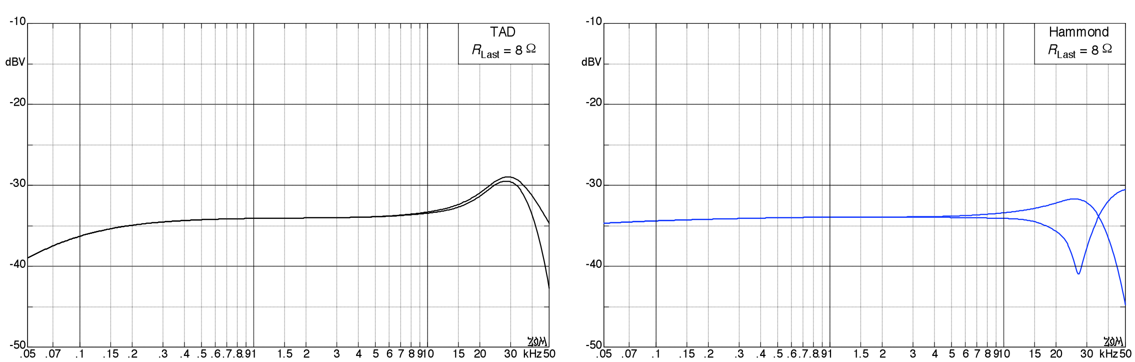


Fig. 10.6.10: Frequency responses of transmission. Primary stiff current-source; asymmetric primary windings.

10.6.4 The non-linear model

Ampère's circuital law describes the connection between the magnetic field-strength *H* and the electric current *I*, while the law of induction characterizes the relation between electric voltage *U* and magnetic flux-density *B*. Both laws are time-invariant mappings. The tie-in between *B* and *H*, however, is given by a non-linear, time-variant mapping: $B = \mu \cdot H$. In the ferromagnetic sheet metals used in transformer cores, the permeability μ is a non-linear quantity the magnitude of which depends both on the field-strength and on past values (compare to Chapter 4).

A first indication of this non-linearity of the core emerges when measuring the transformer impedance. Changing the sinusoidal AC-current flowing through the primary winding of an output transformer, and concurrently measuring the voltage across this winding, we get a quotient depending on the current (**Fig. 10.6.11**). The time-curve of the voltage (or of the magnetic flux-density) indicates strong non-linearity already at moderate amplitudes, i.e. there are deviations from the sinusoidal shape resulting from the warping in the hysteresis-curve (Chapter 4).

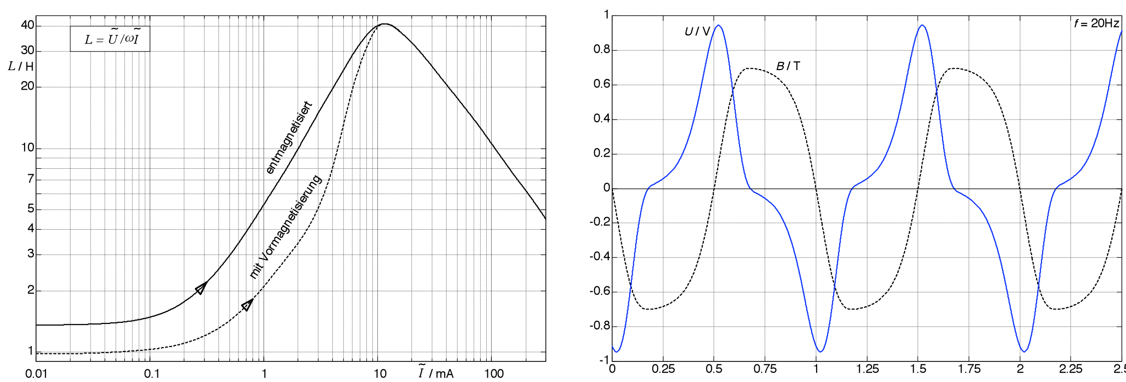


Fig. 10.6.11: Measurement at the primary winding (EI-96). The "inductance" given in the section on the left is a special non-linear quantity. Right: secondary voltage (LL) and flux-density for input from a stiff current source.

The relation between *B* and *H* is, however, not just **non-linear** but in a sense **time-variant**, as well: on the one hand there is an infinite number of hysteresis-loops, on the other hand these can be cycled through only in one direction – for one and the same field-strength there are two corresponding (different!) flux-densities. Of course, the material in the core reacts in the same manner each time if we start from the totally demagnetized state: as such the system is time-invariant. After switching off an external source, however, the core material remains in a partially or fully magnetized state for any length of time, and as we re-start driving the material, an individual characteristic results that is dependent on the previous drive-state – as such there is time-variance. Fig. 10.6.11 includes two curves: the upper was measured with a fully de-magnetized core while the lower resulted from the core having first been strongly magnetized by a DC-field that was switched off for the *L*-measurement – i.e. a degree of magnetization remained (remanence). Last, we need to consider that small drive-states run around an offset-point do not follow the large hysteresis curve (see Chapter 4.10.3, reversible permeability). All these non-linear and time-variant effects give measurements with output transformers a certain challenge. Moreover, the data of the transformers under scrutiny are, as a rule, not known and can be (non-destructively) determined only approximately – the curves shown in the following will therefore include tolerances.

Ferromagnetism is a characteristic of the crystal lattice: the elementary magnets are grouped as Weiss domains, and in demagnetized ferromagnetic materials the orientations in space of these domains are randomized i.e. their combined effects on the outside world cancel each other out. An exterior magnetic field (e.g. caused by an electric current) shifts the borders of the Weiss domains (Bloch walls), and a polarization results. These wall-shifts (in part reversible and in part irreversible) depend in strongly non-linear fashion on the magnetic field-strength – this is the basis for the non-linear electrical behavior. The relation between field-strength H and flux-density B is shown, for small drive-levels, in **Fig. 10.6.12**: it is evident how the hysteresis-loop tilts upright with increasing drive-level, and how consequently the permeability increases. The right-hand picture indicates the field-strengths measured with imprinted flux-density: already at small drive-level a deviation in shape occurs, as does an increasing phase-shift relative to the flux-density curve (dashed line, sketched in without scaling).

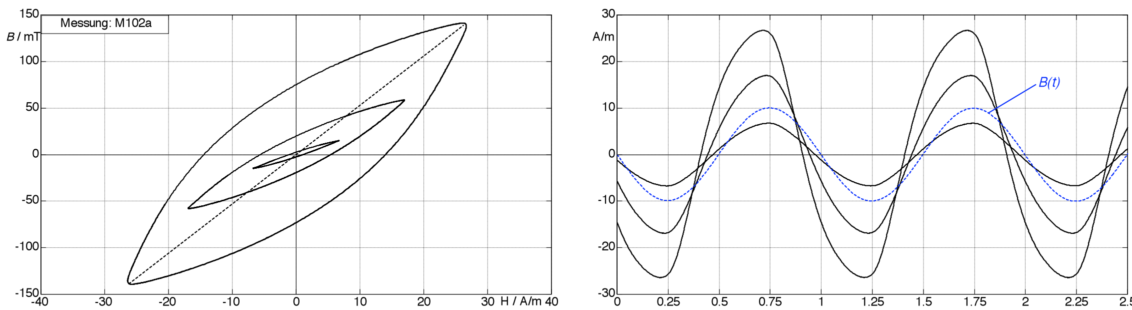


Fig. 10.6.12: Hysteresis loops. Right: time-functions of field-strength measured with imprinted sinusoidal flux-density; dashed: the time-curve of a flux-density (no scaling).

The imprinted flux-density shown in Fig. 10.6.12 is easily achieved: driving a winding from a stiff voltage-source results in an **imprinted flux**♣ (due to the law of induction). In this mode of operation, the voltages transferred to the other windings are also sinusoidal with good approximation – however, this is not the typical case for tube power stages. The latter (as current sources) imprint a priori the current, and this leads to non-linear distortion in the voltages across the windings. This mode of operation is depicted in **Fig. 10.6.13**: already for relatively small field-strengths, non-linear distortion in the flux occurs, leading (as the derivative) to distortion in the voltage. This is not crossover-distortion from the tubes, but pure hysteresis-distortion (imprinting the field-strength works almost distortion-free here).

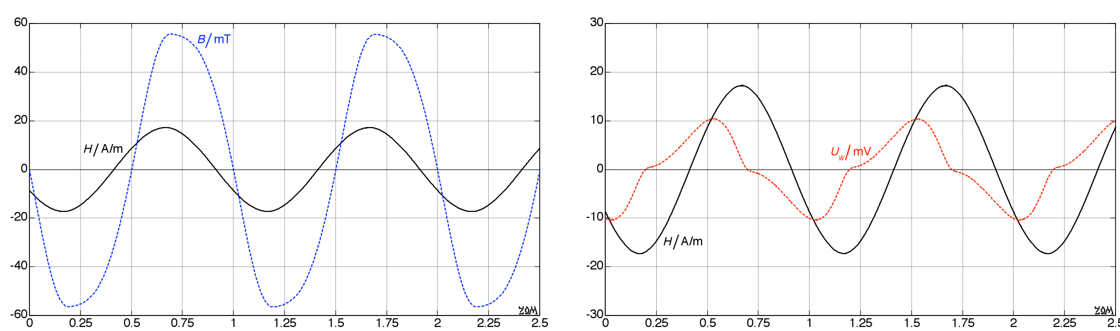


Fig. 10.6.13: Sinusoidal field-strength H (imprinted via the primary current) and corresp. flux-density B (left); non-linear distortion in the voltage U_W across the winding resulting from this H and B (right).

♣ The voltage-drop across the copper-resistance may be compensated, if necessary.

The curves shown in Fig 10.6.13 were measured at an EI-96-core for a secondary open-loop circuit. With a load connected to the secondary winding, this kind of non-linearity increasingly takes a backseat as the frequency rises. If we exclude the transmission of high frequencies for the time being, the equivalent circuit-diagram (Fig. 10.6.3) may be drastically simplified: the secondary copper-resistance R_2 (≈ 0,5 Ω) is added to the nominal loudspeaker resistance, and the leakage-inductance may be omitted, just as the winding-capacitance C_1. The model thus has a purely ohmic secondary loading. Transforming this secondary load via the transformer with TR^2, we get – on the primary side – an equivalent load-impedance $R' = \ddot{u}^2 \cdot (R_2 + R_L)$ connected in parallel to L_1. We may take as guide value for this primary load-impedance about R' = 1 kΩ, as long as we involve *one* primary winding[♣]. Relative to this value, the iron-losses (R_{Fe}) may be neglected, and only three elements remain in the ECD: the primary copper-resistance R', the non-linear parallel inductance L_1, and the transformed load-impedance R' (**Fig. 10.6.14**). The primary current therefore splits up into two parts: the non-linearly distorted magnetizing-current (through L_1), and the current through the load. Compared to the current through the load, the magnetizing current becomes increasingly smaller with rising frequency and looses its significance: the non-linear distortion decreases.

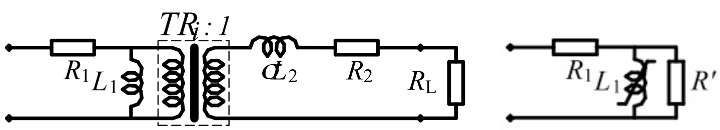


Fig. 10.6.14: Equivalent circuit for the transformer (left); two-pole simplification for low frequencies (right).

It has already been mentioned that this parallel inductance is non-linear; therefore, strictly speaking, no transmission function can be established. The quotient of RMS-source-current and RMS-output-voltage may still be determined, and it is shown in **Fig. 10.6.15** (left-hand section). In the right-hand section, two peculiarities stand out: the slope is not 20dB/decade, and the **cut-off frequency** is drive-level dependent: with increasing drive-level, the low-frequency response improves. As can be seen, it is not purposeful to determine the main inductance based on the initial permeability (as it would be called for according to the classical dimensioning-rule). This approach would land us in the µW-range, which is rather academic in the world of guitar amps. Rather, one could (and should) orient oneself according to the saturation-behavior of the core-material, and determine, for high drive-levels, the flux-density. The saturation of the latter gives hints towards the dominating magnetic distortion.

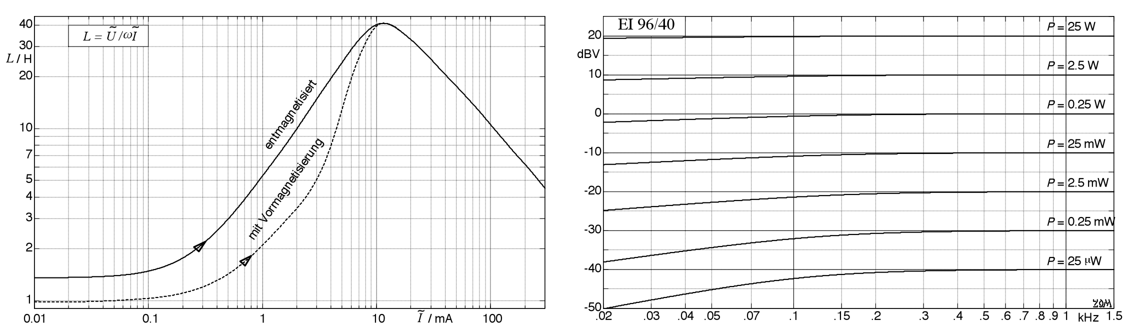


Fig. 10.6.15: Left: drive-dependent main inductance (—— core demagnetized, ----- with remanence).
Right: drive-dependent non-linear high-pass (fed from a stiff-current-source, core of transformer demagnetized). The specified power is fed to the ohmic nominal impedance (4 Ω) at 1 kHz.

♣ For *both* primary windings the quadruple value (not the double) is to be used (Chapter 10.5.5).

Before we occupy ourselves in more detail with the magnet distortions, first a comment regarding the pre-magnetization and **de-magnetization** of the core: we must not expect that the core is always operated free of remanence. At some point, there will be a strong magnetization (even if it happens only as the switching-on impulse occurs), and from this the operating point will return to a point on the hysteresis that does not necessarily correspond to the flux-free origin of the coordinates. Another issue merits attention: only for exactly corresponding plate-currents will the output transformer in push-pull power-stages not experience any pre-magnetization. In most case, the plate-currents will be different, and the resulting difference-current *will* magnetize the core. Consequently, the main inductance will become smaller, and the even-order distortions will increase.

For the demagnetized core (!), the hysteresis loops are point-symmetric, and therefore the distortion spectrum contains only odd-order harmonics. Usually, the 3rd order distortion-suppression a_{k3} is stated; given certain circumstances also the 5th harmonic may be evaluated. The levels of the higher-frequency harmonics are often negligible in comparison. Fig. 10.6.16 shows the 3rd-order distortion-suppression versus the RMS-power (fed to a purely ohmic nominal impedance). In the power-range important for stage-use (over 0.1 W and over 100 Hz), the distortion-suppression remains above 40 dB i.e. the THD remains below 1%. Compared to the distortion generated by a tube power-stage, this is not a dominating effect. Only for lower frequencies and high power output, the transformer distortion rises again steeply – this, however, will usually be outweighed by tube distortion. Of course, the guitarist is at liberty to demand a powerful and distortion-free reproduction of the fundamentals of his/her 7-string guitar. For this scenario, however, a look at loudspeaker-distortion and loudspeaker frequency-responses (Chapter 11) immediately opens the path towards bass-amplifiers and –loudspeakers.

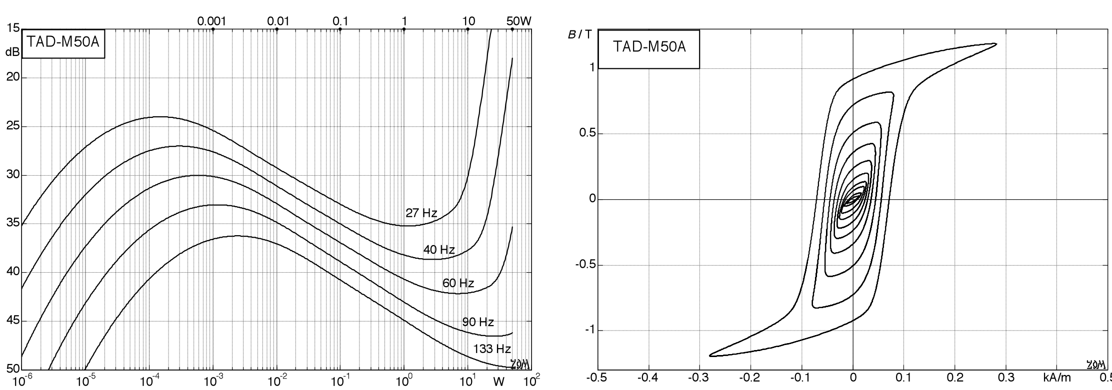


Abb. 10.6.16: Distortion-suppression a_{k3} of a 50W-output-transformer for high-impedance drive-signals and nominal load. The non-linear distortion is generated exclusively by the transformer and not by the driving amplifier. The hysteresis loop shows the relation between magnetic field-strength and flux-density (20 Hz).

A summary in short: the output transformer shows several characteristics that distinguish it from linear, time-invariant components: 1) its main inductance depends on the drive-level; the deep bass is reproduced weaker as the signal level drops. 2) The harmonic distortion is frequency- and drive-level-dependent: the lower the frequency and the higher the signal level, the larger the harmonic distortion; the side-maximum at around 1 mW has little bearing on guitar amplifiers. 3) Harmonic distortion and bass-reproduction depend on the remanence i.e. the previous history of the core-magnetization. 4) How equal (or unequal) the bias-current in the power tubes is, determines the amount of even-numbered distortion components – the matching of the power-tubes is a critical factor here.

The reason for the strange behavior of the output transformer is its warped transmission characteristic. Each of the two transformer windings♣ may be assigned a current and a voltage that are mapped onto each other via transformer and load-impedance. This is classical **systems-theory**: systems map signals onto each other [7]. If a system always reacts the same way, it is time-invariant; if principles of superposition and proportionality hold, and if the system is source-free, it is linear. The transformer is neither – nor. The following considerations concentrate on two (of the four) signal quantities; in a transformer this could be input-current and output-voltage. The nomenclature of mathematical analysis likes to denote the input quantity x and the output quantity y. A so-called "linear function" is defined via $y = 5 \cdot x + 3$. From the point of view of systems-theory, the corresponding system is, however, not linear because "source-free"-condition is not adhered to, among others aspects: in a linear system $y = 0$ has to follow for $x = 0$. A further term needs to be introduced for the consideration of functional dependencies: in a **memory-free** system, the output quantity (y) may, at each and every instant, only depend of the input quantity (x) at that instant. Each pair of values (x_i, y_i) may then be seen as a point on the xy-plane. The entirety of all points forms the graph of the function – this graph is called **transmission characteristic** in systems theory (and it is something completely different from the transmission function). The ideal amplifier features, as transmission characteristic, a straight line traversing the origin. The slope of the straight line is a measure for the amplification factor. The transmission characteristic of the tube (Chapter 10.1.3) is, conversely, bent; the tube therefore amplifies in a non-linear fashion. It is somewhat popular to deduce from this the theorem: "curved transmission characteristics lead to non-linear distortion" – however things are not that simple.

Let us look at the transmission behavior of a simple RC high-pass. Its elements (R and C) are linear components, and therefore the transmission behavior needs to be linear. However, as we plot, for a sinusoidal input-signal, the output quantity versus the input quantity, an **ellipse** (**Fig. 10.6.17**) is generated, i.e. a **curved line**. On top of that, this curve will change shape if the input signal is not sinusoidal anymore. From these simple examples alone, we observe: transmission characteristics are purposeful if the system is memory-free – in dynamic (memory-containing) systems, there is no static transmission characteristic but, if anything, a signal dependent function-graph.

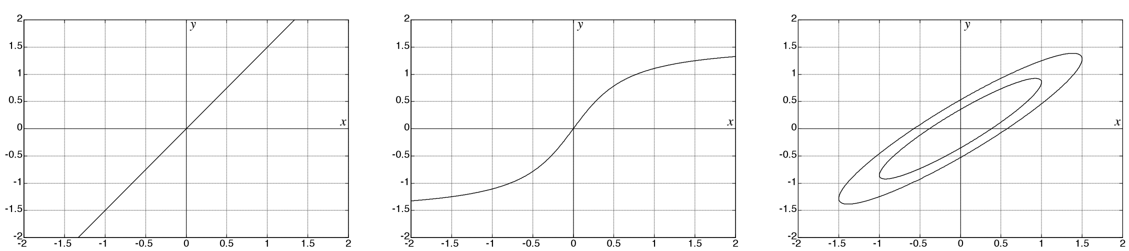


Fig. 10.6.17: Transmission characteristic of a linear system (left) and of a non-linear system (center). For dynamic (memory-containing) systems (right) two drive-levels are depicted.

So, how does that fit with our transformer? Globally viewed, we have a degressive functional relation between magnetic field-strength (abscissa) and magnetic flux-density (ordinate), similar to the curve shown in the middle section of Fig. 10.6.17. In addition, the curve splits into two loop-shaped branches. A family of degressively clinched ellipses is the result (Fig. 10.6.16). Without a doubt this is non-linear, and it is dynamic (memory-including). Still, it is very different compared to the simple RC high-pass.

♣ For the present considerations the primary winding is not subdivided.

The dynamic behavior of the RC high-pass results from recharging processes in the capacitor: after e.g. a step in the input voltage it takes a while until the capacitor has recharged to the new voltage♣. This “while” (i.e. this delay) leads to phase shifts, and these are the reason why the straight line becomes an ellipse. In the ferromagnetic **iron core** of the transformer, the magnetic flux instantly follows the field-strength, any inertia effects (that in fact exist) do not play a role at the very low frequencies considered here. The contoured, s-shaped hysteresis-curve holds for quasi-stationary processes, as well, i.e. for arbitrarily low frequencies.

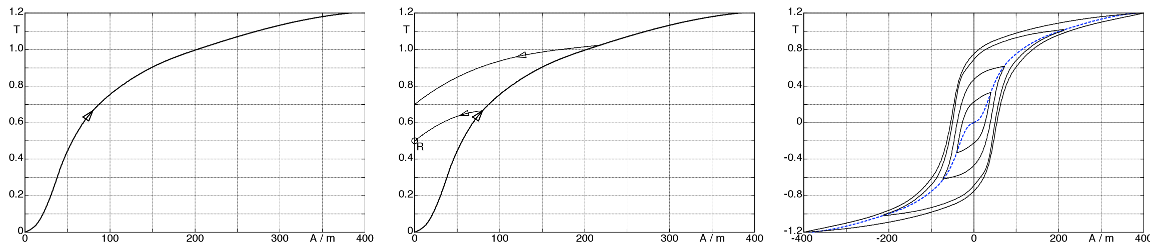


Fig. 10.6.18: Relationship between magnetic field-strength *H* and magnetic flux-density *B*.

The left-hand section of **Fig. 10.6.18** shows the *B/H*-relationship for an initially totally demagnetized core – both *H* and *B* are zero. With increasing field-strength, the flux-density first follows on a progressively bent curve, and on a degressively bent curve. If – starting from any one point – the field-strength is now reduced, the corresponding *B*-value does not wander back along the curve it followed on the upwards path, but it takes a significantly flatter backwards-curve (middle section of the figure). If the field-strength oscillates between two values equal in magnitude, the *BH*-curve encloses the origin, as shown in the right-hand section of the figure for four cases. The quotient of *B* and *H* (the slope of the curve) is proportional to the inductance *L*.

For a very small drive-level, the hysteresis curve has a shallow shape (but is not horizontal), and the inductance is relatively small. In this range, the *B/H*-relationship may be described via two parabolic branches that themselves can be approximated by a flat ellipse (**Fig. 10.6.19**). The parabolas result in a non-linear mapping while the ellipse is linear. As the drive-level increases, the parabolas (or the ellipses) raise themselves up more steeply, and the inductance increases until, at high drive-level, the core material is increasingly saturated, and the slope of the curve becomes flatter again. While this non-linear behavior does not seem to be very complicated, we need to also consider that the loudspeaker-voltage does not depend on the flux-density *B* but on the time-derivative of it ($U \sim dB / dt$). If the drive-signal is not generated by an ideal voltage- or current-source, both voltage and current will be non-linearly distorted and shifted in phase, and on top of this the non-linear inductance is dependent on the drive-level.

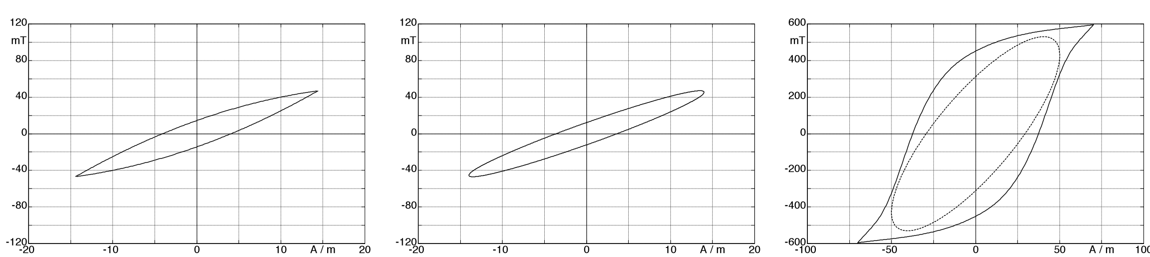


Fig. 10.6.19: Approximations using parabola (right) and ellipse (center). Limits of the ellipse-approximation as saturation sets in (right).

♣ Strictly speaking, it takes infinitely long but we do not need to exactly look into this issue here.

The equivalent circuit diagram developed in Fig. 10.6.14 is helpful to understand these linear and non-linear mappings. For small levels and low frequencies, the main inductance L_1 remains relatively small. For constant output power (e.g. 1 µW), the primary current is (due to $U \sim \omega LI$) inverse to the frequency; in the measurement of the distortion-suppression shown in Fig. 10.6.16, the current-level therefore needs to drop by 3.5 dB while the frequency is increased by a factor of 1.5. Since, as a first approximation, the 3rd-order distortion depends on the drive-signal amplitude according to a square law, the distortion-suppression will correspondingly increase by 7 dB – this can be measured with good accuracy for small power levels (e.g. 1 µW). As the power increases (while the frequency is kept constant), the distortion rises, but at the same time the inductance will, above a certain value of the current, start to increase (Fig. 10.6.15). As soon as the impedance of this growing inductance has reached the size of the transformed load-impedance, the distorted magnetizing current looses significance and the distortion decreases. In Fig. 10.6.16, this is the case at about 1 mW for the 90-Hz-curve. As the power (or, more precisely, the flux-density) continues to increase, the range of non-linear flux-limiting is reached at about 1 T – the distortion suddenly increases. The rather capricious distortion-behavior seen in Fig. 10.6.16 is explained that way, at least as far as the pure transformer-distortion is concerned. It has already been elaborated elsewhere that power tubes and loudspeakers will also operate in a non-linear fashion, and that in particular the loudspeaker impedance may have a strongly non-linear characteristic.

The cause for all non-linear transformer-distortion is found in the non-linear permeability of the **core metal sheets:** it is conducive to examine their magnetic parameters more closely. To guide a magnetic field with low resistance, a material with very high permeability is required: ferromagnetic material with its main ingredient being iron (ferrum). Unfortunately, iron also conducts electrical current relatively well, and for this reason eddy currents can develop their dampening effect at high frequencies without much hindrance (see also Chapter 5.9.2.4). In order to hamper this, a few percent **silicon** are mixed into the iron. Already merely including 1% Si, the electrical conductivity can be halved; it even drops to 1/5th with 5% Si. This is desirable, but the instruction leaflet points to side effects: the saturation limit decreases with increasing Si-content, and the metal becomes more brittle. According to Heck [21], at more than 3.5% Si the metal will break when bent cold, and hot-processed sheets contain 4.5% Si at most. **Fig. 10.6.20** shows **commutation curves** of typical sheet metals for transformer cores. These curves result as the reversal points of the inner hysteresis curves are connected; they correspond practically to curves for previously demagnetized material (dashed in Fig. 10.6.18). Including silicon has a further advantage: the permeability at small drive-levels increases, and the **re-magnetization losses** decrease (Chapter 4.10.4). The main reason that the ideal values presented in the datasheets are not reached in practice is found in the unavoidable **butt joints**: due to the very big difference in permeability between air and core-sheet, even very short air gaps (0.1. mm) deteriorate the magnetic resistance.

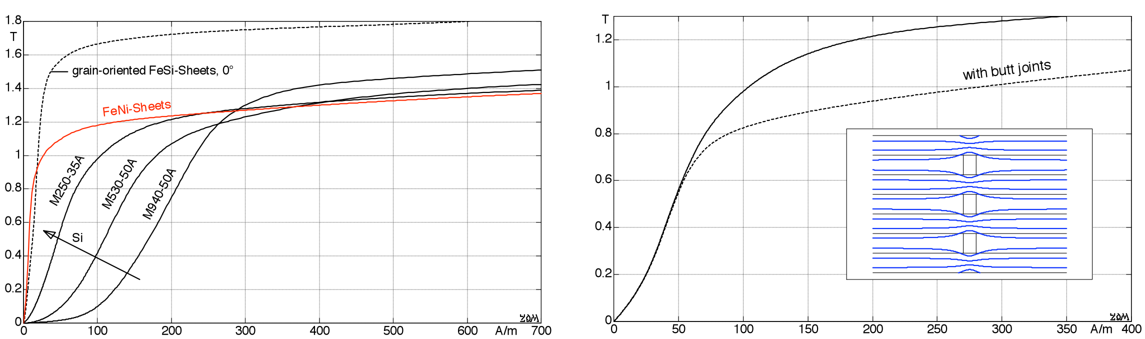


Fig. 10.6.20: Magnetic commutation curves of various core metal sheets; impact of the butt joints.

If the core laminations are reciprocally layered – as it is indicated in Fig. 10.6.20 – there will be 4 overlapped butt joints per magnetic circuit in an EI-core. At each butt joint, the flux density in the neighboring sheet is doubled, and the saturation limit consequently decreases. For the example in the picture, an effective gap-width of 0.2 mm was assumed; the geometric gap-width is even smaller. What's clear here: a sloppy manufacturing process can quickly cancel out any advantage that low-loss core sheets may bring.

How big then are these **core-losses**, anyway? For $\hat{B} = 1\text{T}$, the datasheets specify a power dissipation of 1 – 2 W/kg, i.e. 0.5 – 1 W for your regular 18-W-transformer (500 g_{Fe}). This is for 50 Hz. The often-voiced fear that these re-magnetization losses would rise proportionally with frequency (because the hysteresis loop is traversed more often as the frequency increases) fortunately is incorrect: the *voltage* is approximately constant vs. the frequency♣, and therefore the drive-level decreases with increasing frequency. Besides, if a transformer was to 'loose' 1 W at 50 Hz, it would have to 'loose' 200 W at 10 kHz. No – while these losses do exist (in one transformer somewhat more, in the other somewhat less pronounced), they are not creating any existential danger. It is therefore not necessary, either, to use **NiFe-sheets** with the 20-fold price tag. Already 50 years ago, H. Schröder wrote: *time and again it shows that, for transformers that need to transmit high power, it does not lead anywhere to use materials with high permeability such as permalloy or permenorm. These materials are much too easily overdriven [Lit.].* That's not entirely wrong but requires a supplement: **permalloy** is a NiFe-alloy with 70 – 81 % nickel-content. It allows for very high permeability values but has a rather meager saturated flux density of 0.8 T. In **permenorm** (as mentioned by Schröder), the nickel content is lower (36%) and the saturated flux density higher (1.4 T). These days, 50%-NiFi-alloys reach as much as 1.6 T – almost as good as FeSi-sheets (2 T).

The **saturated flux density** is often connected to the maximum power that can be transmitted – unjustly so in most cases, as the following example will show: the primary winding is connected to a voltage-source, the secondary winding is without load (open circuit), and the primary current mostly depends on the main inductance. We now connect a secondary load-impedance (purely ohmic), and the primary current increases. The smaller the secondary load, the higher the primary current: the more the hysteresis curve is pushed? Given Ampère's law, isn't that correct? In fact, it isn't: the now flowing higher secondary current generates a magnetic field, as well, and this one is oriented in the opposite direction of the primary field (Chapter 10.7.6). The core-drive depends on: voltage, frequency, and inductance $\Phi \sim U/\omega L$. In the power stage, the maximum amplitude of the voltage is determined by power supply, and by the tubes – it is, as a first approximation, constant. Given this, and a specific frequency (e.g. 100 Hz), the drive-level in the core is halved as the permeability is doubled. Relative to FeSi sheet metals, datasheets specify a 10 – 20-fold higher permeability for NiFe-sheets – a slightly smaller maximum flux density would not be of any bother here, would it? Indeed it wouldn't – if the core actually had such a high permeability. However, the larger the permeability of the material, the more the unavoidable air gaps make themselves felt. NiFe sheet metals are therefore purposeful predominantly for tape-cores. According to Boll, EI-cores are almost exclusively fabricated from FeSi-sheets, and M-cores in small number from NiFe-sheets. In the end, an optimization is required that considers, apart from permeability and saturation flux density, also iron-losses, build-size and – especially – cost. Whether a core costs 7 Euro or 100 Euro is crucial. If there is too much distortion, a slightly larger FeSi-core should also be considered (instead of the NiFe-core). It would be far less pricey. At the time of this writing (2012), sheet metals with high nickel content cost about 60 Euro per kg – given a minimum purchase of 50 kg.

♣ It's not perfectly independent of frequency, but $U \sim f$ certainly does not hold, either.

Another alternative approach that may be taken is found in **grain-oriented transformer sheets**. Applying special milling and annealing, these sheet metals receive a preferred orientation (**texture**); they are **anisotropic**. In a specified direction, their permeability is higher than that in isotropic SiFe-sheets, and the re-magnetization losses are correspondingly smaller. In tape-wound cores and split-tape cores, this advantage takes full effect. In EI- and M-cores, the additional price needs to be carefully weighed against the quality-increase because here the magnetic flux will in places run transversely to the preferred orientation. **Fig. 10.6.21** contrasts hysteresis curves as published by the manufacturer of base-materials with measured curves. The shapes do not match exactly for a number of reasons: 1) stamping will deteriorate the material properties at the stamping-edges; 2) The butt joints (unavoidable in EI-cores) decrease the maximum magnetic flux; 3) with grain-oriented sheets (M165-35S), the flux is oriented in unfavorable directions also, e.g. transverse to the preferred orientation. It is rather striking here that the data of the base-materials are not achieved.

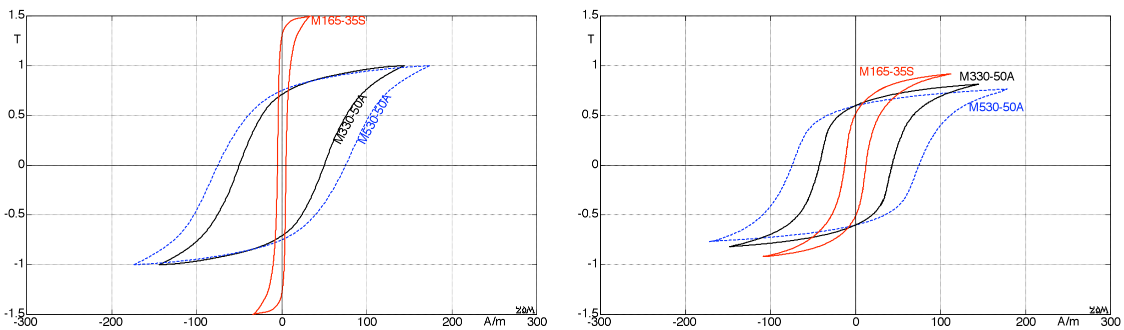


Fig. 10.6.21: Material characteristics (Waasner, left), measurements (EI96a, right). The material characteristics are valid for the base-materials; stamping will change the values; for the influence of butt joints: see Fig. 10.6.20.

Fig. 10.6.22 shows how big the **orientation dependency** in grain-oriented transformer sheets is: at an angle of 60° and 90° we obtain curves as they would result for regular, non-grain-oriented sheet metal. It is consequently not surprising that the good values featured by the base material are not achievable with EI-cores – even with meticulous assembly. All too easily the impression could be created that the air-gap between the E and I of an EI-core (Fig. 10.7.14) could be avoided if both these sheets were only pressed together tightly enough. However, these are non-planar, non-parallel surfaces that meet. The boundary surfaces result from stamping, and they are slightly arched such that even with peak compression, gaps remain. The datasheets have info about which tolerances are desirable: 5 µm are seen as good quality; this is a value that cannot be achieved with stuck-together EI-sheets. Even for split-tape cores, this could only be obtained with optimum bracing – the long-term sustaining of which is not at all trivial.

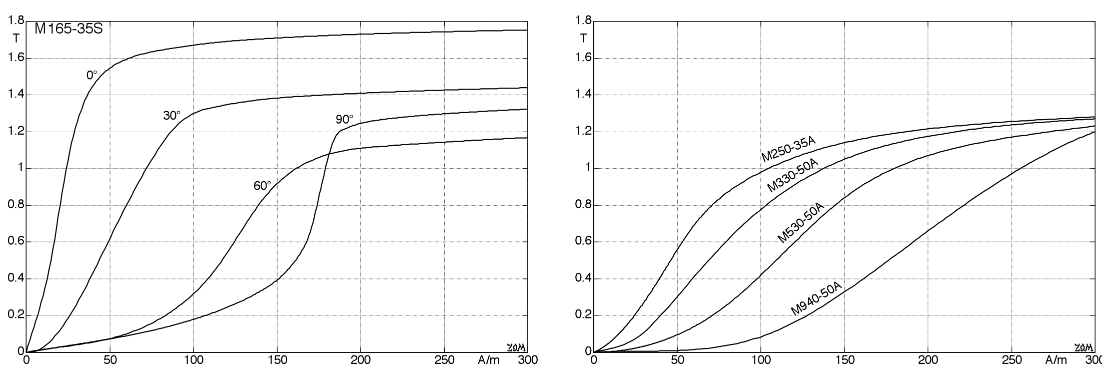


Fig. 10.6.22: Magnetization curves: grain-oriented sheets (left), isotropic sheets (right); base material.

We manufactured three transformers using the three core sheets mentioned above with 900 turns on the primary winding and 79 turns each on the respective two secondary windings♣. The **inductance** ((*U/ωI*, measured via the primary RMS-current) is shown in **Fig. 10.6.23**: although the grain-oriented sheet metal does not reach the nominal data of the base-material, it still clearly outperforms the isotropic sheets. It is, however, also significantly more expensive. As an effect of the enlarged inductance, we obtain a smaller **harmonic distortion**, as depicted in the right–hand part of the figure. In the budget-priced M530-50A, and at 80 Hz and 50 W, the THD is four times that found in the M165-35S. Before we elect a favorite, though, it is wise to take a look at Chapter 11.6: the non-linearity of regular guitar loudspeakers is much higher that that of the transformers examined here.

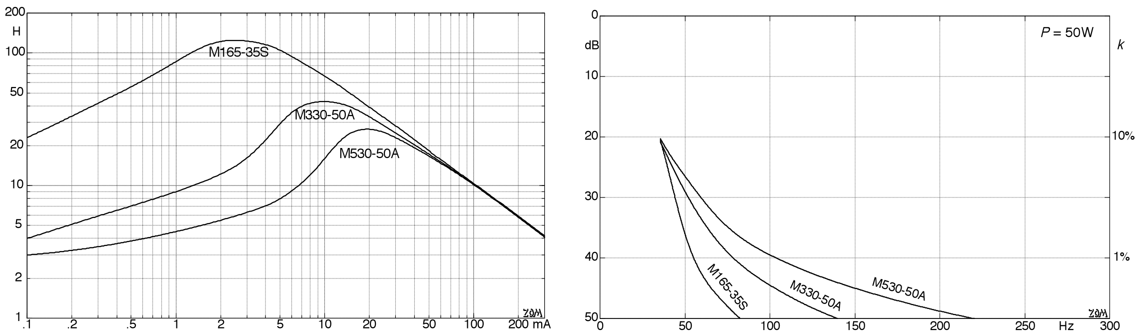
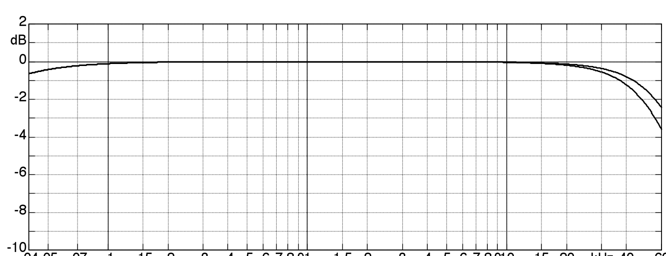


Fig. 10.6.23: Inductance of *one* primary winding (N = 900, RMS current); distortion-suppression at P = 50 W. M330-50A and M530-50A are isotropic FeSi-sheets, M165-35S is a grain-oriented FeSi-sheet. EI-96a.

Besides the harmonic distortion, the **frequency response** is of course also of interest – the windings♣ were not nested, after all – so according to popular HiFi-lore no usable outcome could be expected. **Fig. 10.6.24** shows, however, how viable the result turned out to be. The transformer was connected to a secondary load of 8 Ω, and for each measurement one of the two primary windings was driven via an internal impedance of 8 kΩ. Nesting the windings will drive up cost, and make the filling factor of the copper drop. The Cu-resistances of the transformer investigated here are R_{aa} = 53 Ω, and 0.17 Ω for the 8-Ω-winding. This is not bad at all, compared to the industrial products examined in Chapter 10.6.5, the **Cu-resistances** of which are two to three times as high, with correspondingly higher thermal copper-losses. The **iron losses** cause few problems: for the investigated EI96-transformers, we found as little as 1.2 W (M350) and 0.55 W (M165) at 1 kHz and 50 W. As expected, the grain-oriented sheets win out – but the advantage is, absolutely taken, insignificant. Simple **conclusion**: in a guitar amplifier, expensive core sheets have a hard time pushing their advantages. **The M330-sheet represents a good compromise.**



Primary winding W1b
Secondary winding W2b
Secondary winding W2a
Primary winding W1a

Fig. 10.6.24: Frequency response for an 8-Ω-load. Primary drive via 8 kΩ, P = 1/4 W.
Both secondary windings (1 mm ∅) are connected in parallel, EI-96a core, core sheet M165-35S.

♣ Since no 1,5-mm-wire was at hand, 2 secondary windings were set up using 1-mm-wire.

10.6.5 Comparison measurements

From 2012 to 2016, the university at Regensburg (Germany) offered a practical course on tube amplifiers for which a modular guitar amplifier was developed. It included a 15-W-power-stage with the possibility to directly switch between up to 10 different output transformers, and a 50-W-power-amp offering a choice between 13 OT's. The candidates are:

	Transformer	Z_{aa} / kΩ	$R_{aa=}$ / Ω	$R_{8\Omega=}$ / Ω	Core	Amplifier	€
1	Conrad ELA 10W	7,9	280	0,70	EI-48/16	Ela	6,90
2	Conrad ELA 20 W	7,0	180	0,33	EI-48/24	Ela	9,50
3	Hammond-1750E	8,8	300	0,45	EI-57/19	Deluxe Tweed	34,70
4	TAD-1839	9,1	560	0,70	EI-66/22	Deluxe Tweed	86,20
5	TAD-125A1A	6,9	330	0,44	EI-66/22	Deluxe Reverb	69,00
6	Hammond-1760H	5,9	400	0,83	EI-66/22	Deluxe 'upgrade'	54,39
7	Hammond-1750J	8,2	180	0,35	EI-75/24	Tremolux	38,65
8	TAD-MJTM18WA	9,1	670	0,60	EI-75/24	Marshall 18Watt	79,00
9	Hammond-1750Y	6,8	300	0,50	EI-75/38	VOX AC15	77,30
10	NSC 401318-T	7,1	196	0,50	EI-66/22	e.g. Fender	17,80
11	TT-SLO50	4,5	100	0,43	EI-96/40	Soldano 50W	88,90
12	Hammond-1760L	4,1	100	0,41	EI-96/31	Bassman 'upgrade'	82,30
13	Marshall JTM-50	3,5	86	0,54	EI-96/40	Marshall 50W	86,56
14	Hammond-1750N	3,5	80	0,51	EI-96/40	JCM800	77,50
15	OTH M330-50A	3,5	53	0,17	EI-96/36	university lab	--
16	Hammond-1750V	4,2	140	0,70	EI-96/40	VOX AC30	86,50
17	Hammond-1750Q	**7,9**	140	0,61	EI-96/40	JTM-45	92,25
18	Marshall JTM-45	**7,8**	155	0,42	EI-96/40	JTM-45	100,30
19	IG-Wickeltechnik	**8,2**	218	0,49	EI-96/40	JTM-45	106,20
20	Toroid mains transf.	3,5	60	0,21	∅81x35	Mains transformer	15,--
21	TAD-MJTM45A	**8,1**	360	0,49	EI-96/40	JTM-45	129,50
22	TAD-018343	4,7	100	0,20	EI-96/34	Super Reverb	110,00
23	TAD-M50A	3,7	150	0,48	EI-96/40	Marshall 50W	89,90

The 'small' transformers (upper group) are operated at either 2xEL84, or 2x6V6-GC while the 'big' ones work with either 2xEL34, or 2x6L6-GC, or 2xKT-66. Using the easily accessible datasheets as a basis, the **optimum load-impedance** (plate-to-plate, Z_{aa}) across the entire primary winding should amount to **8 kΩ** for both the 2xEL84- and the 2x6V6-GT-complement. Checking a bit more thoroughly, we find as a boundary condition e.g. for the 6V6-GC: a plate- and screen-grid-voltage of 285 V. However, the Deluxe in fact was operated already in its initial versions at 350 V, and later with as much as 420 V. This slight ☺ overload has not killed it (the datasheet allow for a maximum of U_a = 315 V) … but what about the optimum load-impedance at these voltages? The datasheets are silent about it – presumably because of the limit value mentioned above. These days, transformers produced for these amps mostly have about 8 kΩ for the early Deluxe-variants and 6.7 kΩ for the later ones. The measurements in the table indicate that these target specifications are 'generously' interpreted. For the 'big' amps there is agreement that the correct load-impedance for a JTM-45 should be exactly 8000 Ω … that does not prevent TAD to include a 3,7-kΩ-transformer with the JTM-45-kit. Well, you are free to reorder the 8-kΩ-varaint for an extra 130 Euro. Over to the 2x6L6-GC or 2xEL34: here, impedance-values of around 4 kΩ are customary, and you are in good hands with this for the AC30 (4xEL84), as well. It is recommended to take the impedance specifications with a pinch of salt – they are frequency-dependent, and the tube data that are supposed to be a match to these impedance values scatter rather strongly, too.

Fig. 10.6.25 shows the measured frequency responses of the impedance. The transformers were loaded at their 8-Ω-output with 8 Ω, and primary impedance of the *entire* winding (Z_{aa}) was measured. The Tremolux-OT (7) is actually specified for 4 kΩ / 4 Ω, and it was tested with 8 Ω at its 4-Ω-output, which approximately doubles the primary impedance. The two ELA-transformers were not actually specified for operation with a push-pull power stage but their windings allow for comparable transmission ratios. Still, it needs to be emphasized that these transformers were designed for an operation with 100 V and not for 250 V as it regularly occurs with power stages (U_a, under regular operation). Corresponding experiments therefore require adequate safeguarding. All measurements were taken with very small power such that, for the low-frequency impedance, the **initial permeability** is significant. The latter is particularly small for the ELA-transformers; but this was to be expected in the face of the very small build-size. Also, it must not be forgotten that the other transformers are about 10 times the price! The impedance increase at high frequencies is due to winding-resonances and –capacitances, and the scatter in the middle frequency-range is due to differences in the transformation ratio (turns-ratio).

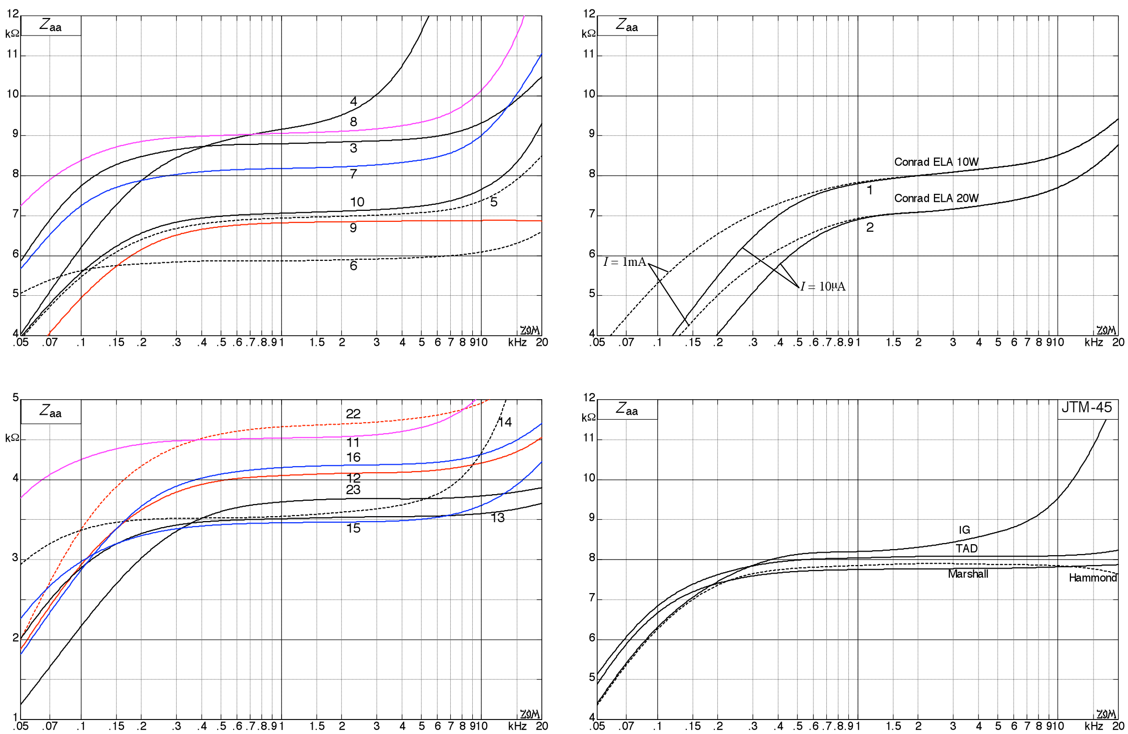


Fig. 10.6.25: Frequency response of the impedance (Z_{aa}) for drive from a stiff current-source (10 µA) and a secondary load of 8 Ω. The numbers in the figure relate to the above table.

It is not imperative to assume that the different transformation ratios result from bad manufacturing quality. The number of the turns of the wire can easily and precisely be checked; divergences are, with high probability, intentional. The suppliers indicate e.g. Z_{aa} = 8.1 kΩ, but apparently a result of 9 kΩ will not be the end of the world. What seems to be more important: *manufactured according to the original specs using authentic materials*. That's the reason for the high price. For the 5E3-Tweed-Deluxe, you will find a vast variety of output transformers; these all wait to be lovingly assembled by hand (and with authentic materials) first, and that costs. One single variant for all 18-W-amps would probably also do – but only for the very un-emotional customer.

Translated by Tilmann Zwicker

Because a purely ohmic 8-Ω-load is required but not sufficient, the corresponding figures with loading by a loudspeaker are also included (**Fig. 10.6.26**). As already elaborated in Chapter 10.5.8, the (*straight*) load line is a first approach – reality is more complex (in the true sense of the term). The power tube does not "see" a constant resistance but a complex load the magnitude of which varies between e.g. 7 and 30 kΩ. This could as well be a range from 9 to 50 kΩ – or whatever else the transformer offers as a load. Depending on the transformer and the loudspeaker, the optimum operational range of the amplifier therefore resides within different frequency ranges, and consequently, the output transformer influences the sound. Again: this ain't no secret science: with the turns numbers, and the size of the core, you have the main ingredients already on the table.

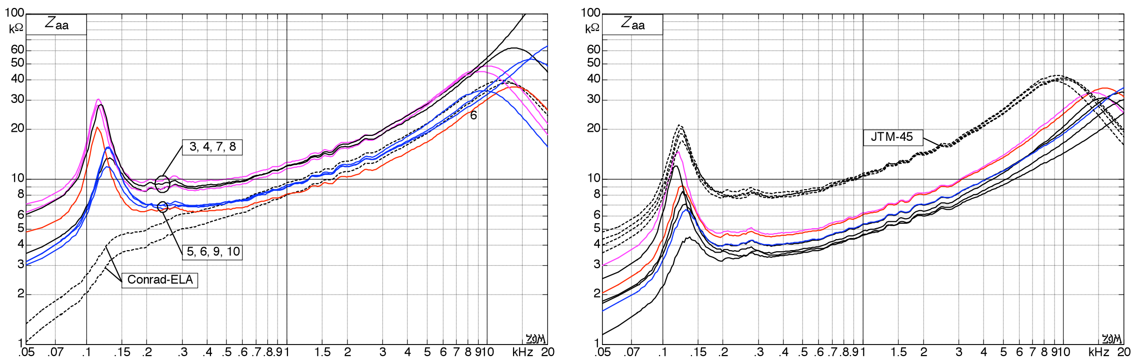


Fig. 10.6.26: Frequency responses of the impedance (Z_{aa}) with drive from a stiff current source (10 μA), load = Jensen C12N in an enclosure. Left: first-group transformers (1 – 10); dashed = Conrad-ELA-transformer. Right: 50-W-OT's of the second group; dashed = JTM-45-transformer (8 kΩ).

The impedance graphs give an impression of the strain on the tubes; more important, however, is the power transmission (**Fig. 10.6.27**). At small output power (0.2 W / 1 kHz), the variation within the transformers is not that big anymore; even the ELA-transformers provide sufficient bass-reproduction. In the range of the power-limit (right-hand section of the figure), however, differences show up, after all. No. 6 (Hammond 1760-H) has the smallest primary impedance (5.9 kΩ) and therefore delivers the highest output power in the frequency ranges where the loudspeaker is of high impedance. The opposite is represented by No. 4 (the TAD Tweed-Deluxe-transformer): its forte is in the area of low speaker impedance i.e. in the middle frequency-range. At small and medium output power, the sound can be shaped via filters almost at will. However, if the power stage is operated in the range of its power limit, we find: **for a brilliant sound, the output transformer should show low primary impedance, and for a more mid-range-y sound, it should feature higher impedance**.

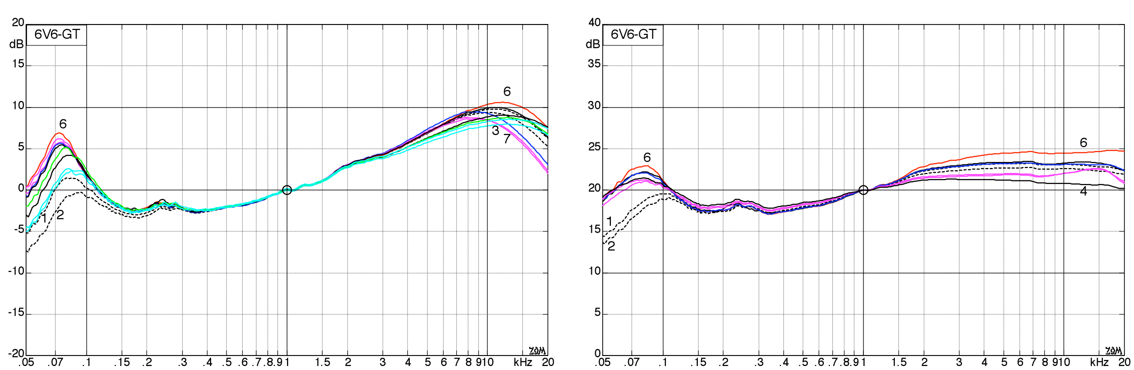


Fig. 10.6.27: Transmission from the phase-inverter input (NFB disabled) to the loudspeaker (P12N).
Normalized to 1 kHz, small drive-level (left), high drive-level with power-stage overdrive (right).

In **Fig. 10.6.28** (not normalized), the effect of the transformer establishes itself clearly; No. 6, with small primary impedance conversely generates the largest secondary source-impedance. This is why the loudspeaker impedance maps itself relatively strongly onto the transmission frequency-response. If the internal impedance of the power stage were zero (ideal current source), the figure would show a horizontal straight line. Relative to this theoretical "ideal" situation (that for a guitar amp would generally be held as not ideal), the source impedances in the figure increase with the sequence 4-3-9-6. The compression of the curves follows almost the same sequence; it is only No. 4 that gets out of line: the TAD-transformer offered for the Tweed-Deluxe (4) has the highest DC-resistance and therefore somewhat higher copper losses. To compensate, it is the most expensive one of them all. And who knows: maybe it is the most authentic one, as well.

Regarding the **strain on the power tubes**, the following holds: the higher the primary impedance, the more the screen grid is likely to be overloaded (Chapter 10.5.9). Thus, if you run your 4-Ω-amp into a 16-Ω-speaker, better keep a watchful eye on your power tubes.

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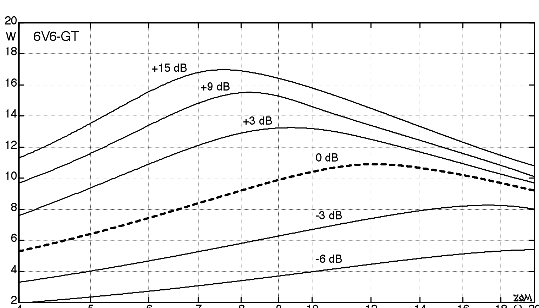
Fig. 10.6.28: Transmission from phase-inverter input (NFB disabled) to loudspeaker (P12N).
This figure is reserved for the printed version of this book.

Let us take another look at the differences between linear and non-linear operation. Between power tube and loudspeaker, there is no tone-stack – if the power tube is clipping, only the output transformer is left to have any impact on the transmission behavior. Therefore, the transformer-ratio is important to the sound. The tube power stage has a relatively high output impedance. If it were as small as it is in a transistor power stage, the output power would increase as the load-impedance decreases. Conversely, the output power increases, in a tube amp, as the load-impedance increases. This will not be the case without limit, though – at some point, the tube hits its limit and then the situation reverses. We see this in **Fig. 10.6.29** for the Tremolux-transformer (7), while Fig. 10.6.30 gives on overview over the remaining measurement results.



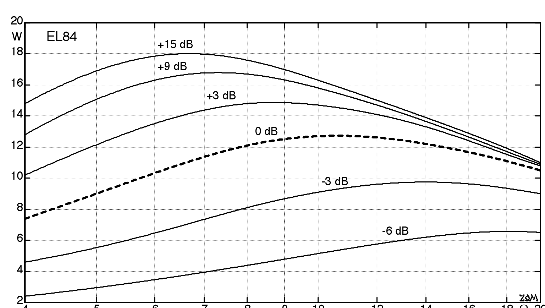


Fig. 10.6.29: Output power dependent on the load-impedance for various drive-levels. At the dashed line, the power-stage overdrive starts. Power stage without negative feedback. 1 kHz.

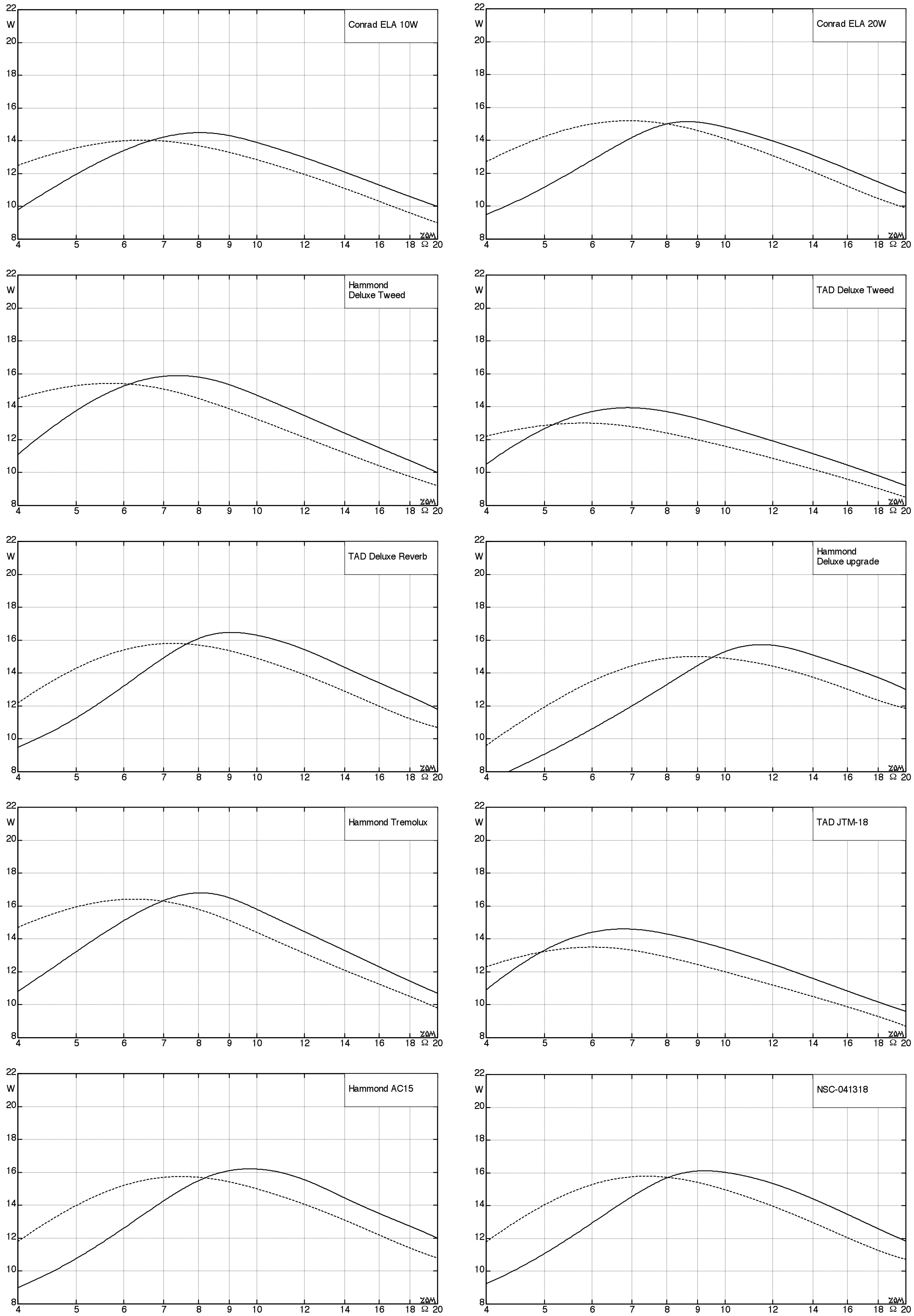


Fig. 10.6.30a: Maximum power vs. (ohmic) load-impedance; power stage overdriven by 14 dB, 1000 Hz. Two different 6V6-GC pairs: Ultron (——), TAD (-----); R_K = 270 Ω // 250 µF.

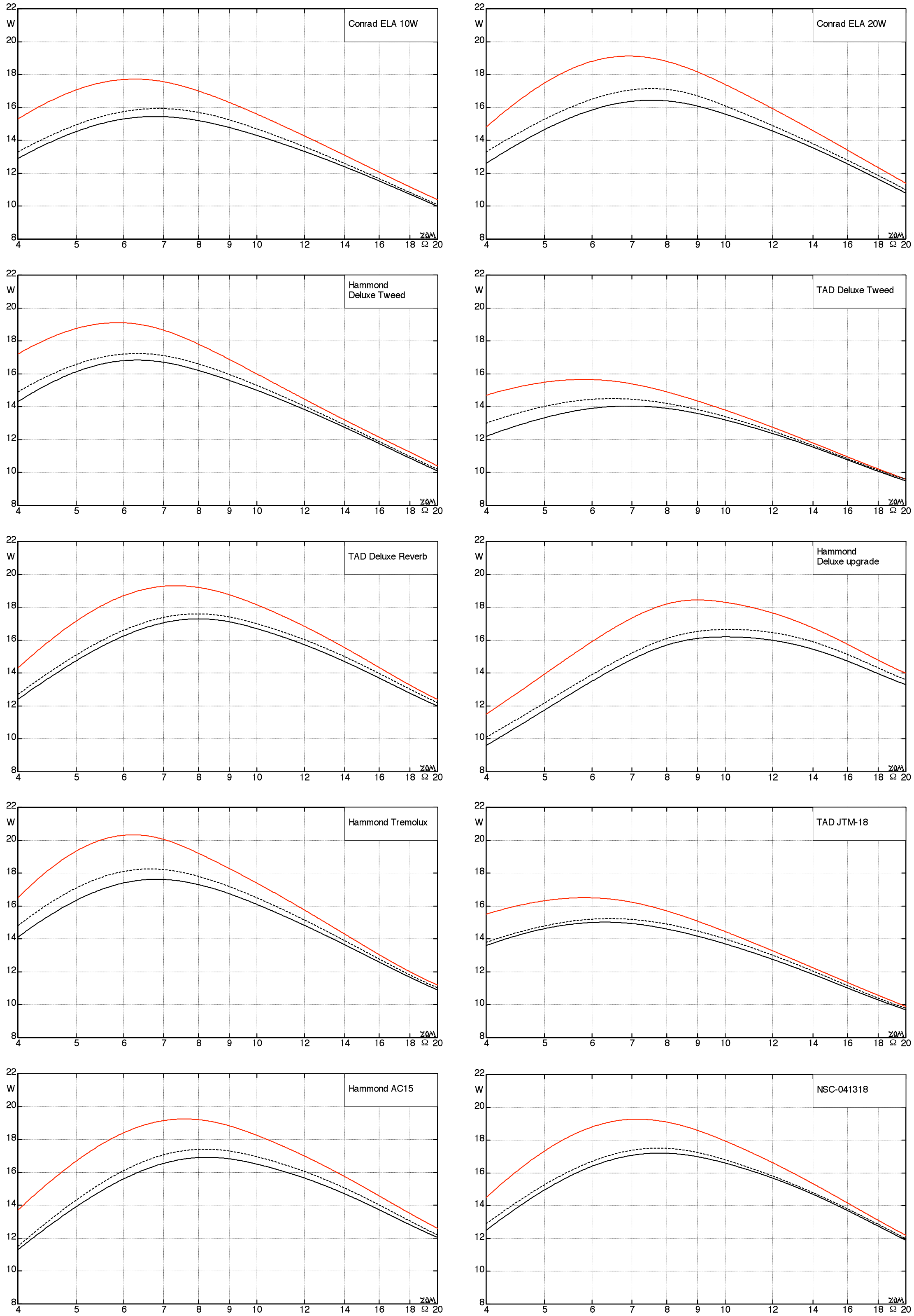


Fig. 10.6.30b: Maximum power vs. (ohmic) load-impedance; power stage overdriven by 14 dB, 1000 Hz. Three different EL-84 pairs: JJ (upper curve), Ultron (------), TAD (lower curve); R_K = 120 Ω // 250 µF.

The preceding diagrams indicated that output transformers may result in different operational behavior – even if offered for the same amplifier model. One parameter in this context is the frequency response under full load (Fig. 10.6.28), another is the maximum power (Fig. 10.6.30), and a third is the **harmonic distortion**. The Hammond transformer investigated in these measurements works (in conjunction with Ultron 6V6-G) most efficiently at a load of 12 Ω. This result is documented in **Fig. 10.6.31**, as well. However, using other tubes, different values were obtained which again shows that the cooperation of several components determines the transmission behavior of the power stage.

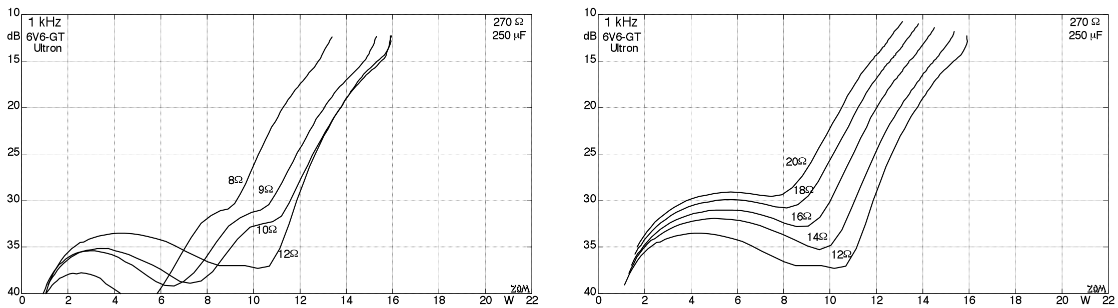


Fig. 10.6.31: Power stage: distortion-suppression for different load-impedances: Hammond Deluxe upgrade. A distortion-suppression of 20 dB corresponds to a harmonic distortion of k = 10%.

Fig. 10.6.32 shows the measurement results for two different tubes (6V6-GT, EL-84) and two different cathode circuits. **EL-84** with R_K is typical for the VOX AC-15 and the 18-W-Marshall (Model 1958). **EL-84** without R_K reflects the Mesa/Boogie Studio-22, and **6V6-GT** with R_K corresponds to e.g. the Tweed Deluxe. **6V6-GT** without R_K is exemplified in the Deluxe Reverb. Because of their differing turns-ratios, the transformers exert a different load onto the power tubes and generate different distortion-suppression that way. This is for 1 kHz, though! Here, another parameter enters the scene: **the frequency**. This now is the point where the depictions start to become confusing. It shall be mentioned only in passing that on top of everything, the plate-voltage, the screen-grid resistor, and the phase-inverter may also vary.

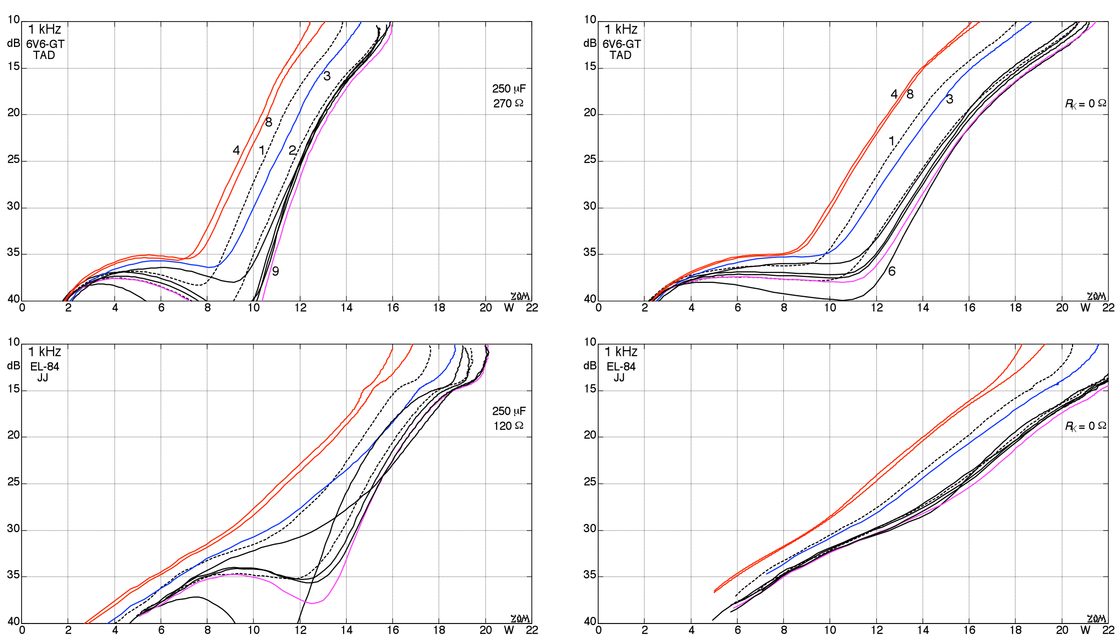


Fig. 10.6.32: Distortion-suppression: power stage with different output transformers. On the respective upper right the cathode circuit is indicated (common R_K bridged by 250-µF-capacitor, and fixed bias, respectively). 8 Ω.

In all these transformer-measurements, only the turns-ratio has shown itself as relevant so far. However, distortion measurements in the **low-frequency** range redirect the attention to the **main inductance,** or the core-material and –size. If the turns-number is too small, the bass-reproduction becomes weak and distorted. An increase in the turns-number, however, can only be achieved (due to the limited space for the winding) by reducing the diameter of the wire, in turn increasing the copper-resistance. If the secondary copper-resistance amounts to 0.83 Ω (as it is the case in the Hammond 1760H), 10% of the generated power remains in the secondary winding. Approximately the same percentage will again be dissipated in the primary winding. If both high efficiency *and* good bass-response are the objective, only changing to a better core will help, resulting in higher weight and/or price. No magic here: the small Conrad-ELA-transformer (1) features merely a cross-section of the iron of 2.4 cm^2 in its small core, and no attention was given to achieving a minimum air gap, either. The result can be seen in **Fig. 10.6.33**: very strong distortion in the bass. With its proud 8.7 cm^2, the AC-15-transformer of course has a much easier life here. It is no contradiction that the 20-W-transformer (2) is even worse than transformer (1): (2) is of particularly low impedance and therefore has an even smaller *L*. Again: these are ELA-transformers!

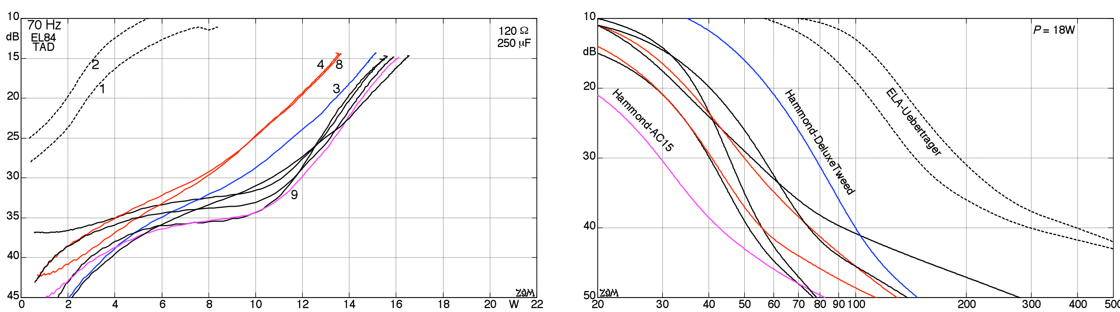


Fig. 10.6.33: Left: distortion-suppression in the power stage for different OT's. 8-Ω-load at the 8-Ω-output. Right: distortion-suppression of the OT (without power stage) as a function of frequency.

The distortion shown in the left-hand section of Fig. 10.6.33 is generated in part by the output transformer and in part by the power tubes, while the distortion shown in the right-hand section stems from the output transformer only. The ELA-transformers experience strong overdrive at low frequencies and are good for distortion sounds, if anything at all. Their primary impedance in the mid-frequency range is a good match for the tubes but their inductance is too small. However, all other transformers are suitable for guitar amplifiers, whether they cost 18 or 86 Euro. Unless blatant errors are made, the following theorem holds: **in the frequency range important for the electric guitar, the turns-ratio (i.e. the primary impedance Z_{aa}) is the decisive parameter; everything else is of minor importance**. Indeed, the manufacturers do use different core sheets, and, yes, they do invest much time in "authentic" replicas. They procure old (i.e. outdated) insulation paper, search for wire insulated in an antiquated fashion, copy scary nesting for the winding, and of course they need to be royally remunerated for the whole hoopla – it is, after all, almost one-off production. **Mindless reproduction of outdated technology on the basis of misunderstood context?** Yes, for the odd transformer this impression does force itself. However, let's not take such a narrow view. Maybe we should consider the approach of the placebo-pharmacologist: where there's a will, there's a market. So: at the latest as we have outgrown our 18-W-shoes and have dragged our 30-W-whopper to the stage despite the slipped disc, we sigh contently: two really fat transformers, thus really fat sound.

We have given the 'small' 18-W-transformers a lot of space – almost too much since more diagrams do not necessarily mean more clarity. Therefore, a short description shall suffice for the 'big' **50-W-transformers**. The frequency responses have already been shown – maximum power and distortion have similar characteristics as with the 18-W-OT's, just with a higher power level. Overall, the quality is somewhat higher, because for the larger transformers the inductance-determining relationship iron-surface-to-iron-length is more favorable. All 50-W-transformers investigated here perform well, whether they have 3.5 kΩ or 4.7 kΩ (R_{aa} each). Supplementing Fig. 10.6.26, **Fig. 10.6.34** depicts the transmission frequency responses of the complete power stage employing EL34's. It is clear that, in the frequency range important for the electric guitar, all transformers work almost equally well♣.

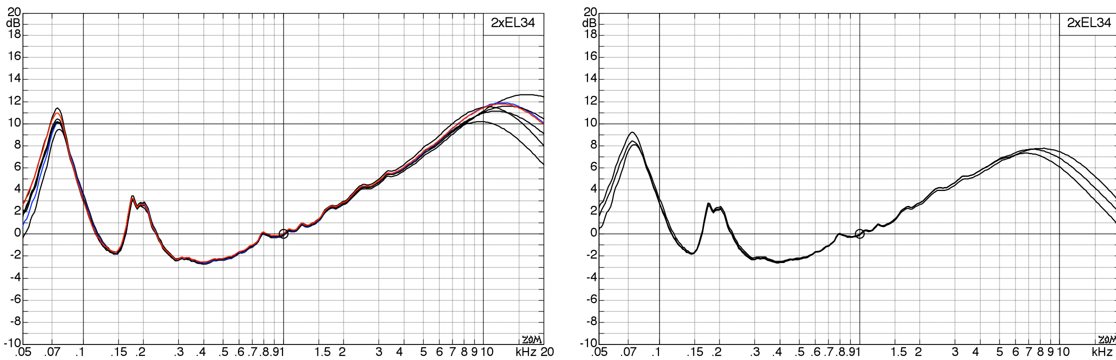


Fig. 10.6.34: Transmission from the phase-inverter input to the loudspeaker (Vintage-30 in enclosure). Right: transformers No. 11 – 16; left: transformers No. 17 – 19.

We see larger differences for the non-linear distortion (**Fig. 10.6.35**). As will be elaborated later, the 8-kΩ-transformer is unsuitable for a power stage deploying EL34's. All other transformers show a similar behavior at and above 90 Hz; it takes a backseat compared to the effect of the tubes. The self-wound M330-50A (compare to Fig. 10.6.23) was a first foray into building an OT – it is suitable, as well. With the addition of 10% more turns, this transformer could have been brought into the range of the other transformers (there would be enough space even with the same wire diameter) – however, this step was not deemed necessary. The red curve refers to a very special "output transformer": a **mains transformer**. Indeed, this works, as well! Not with just any mains transformer – we needed to look around a bit, but this one fits the bill. It's a toroidal transformer costing all of **15 Euro** – is smaller, more efficient, lighter by 1.5 kg, and much less expensive (due to large-scale manufacture). Why do we then still need an EI96? Maybe because it has been done like that for more than 60 years? And because micro-entrepreneurs do like to make the odd 100 – 300 Euro …

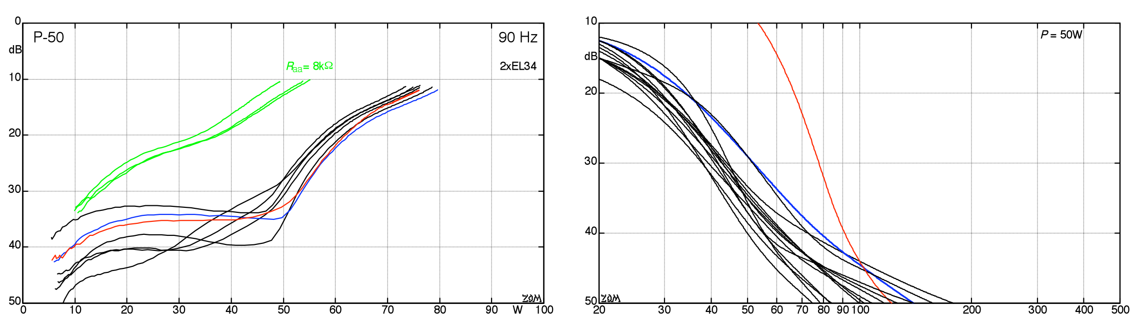


Fig. 10.6.35: Distortion-suppression. Left: whole power stage with 10 different OT's; right: OT's only; *blue*: M30-50A; *red*: mains transformer as OT.

♣ The difference 3.5 kΩ vs. 8 kΩ will be discussed later.

Mains transformers are optimized to achieve optimum efficiency – exotic issues such as "harmonic distortion" are of zero interest in this area. You just wind thick wire and decently drive the core in order to find a good compromise between power-losses and cost. The copper-resistance of the 9-V-winding – which we abused as 8-Ω-winding – reads only 0.2 Ω, compared to 0.4 to 0.7 Ω for the real OT's. Given a few more turns, the main inductance could be increased without significant deterioration (and especially with next to no additional cost), and the harmonic distortion could correspondingly be lowered. This is not intended as a general call: "guys 'n' gals, just load your power stage with a low-cost mains transformer", but it means to say that, given the correct calculation, and fabricated in industry-correct quantities, a toroidal transformer can be a small, light-weight and inexpensive alternative. And what about the frequency response? It's fully in the green, as shown by Fig. 10.6.34. For the sake of completeness, the measurements with an ohmic load are shown in **Fig. 10.6.36**. All OT's are perfect – including the mains transformer..

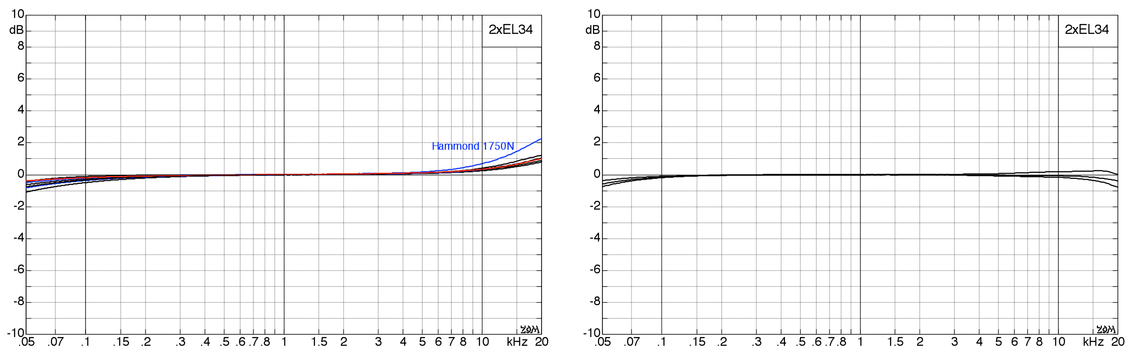


Fig. 10.6.36: Frequency response at an ohmic load (8 Ω); right: 8-kΩ-transformers. On the left, the frequency response of the mains transformer (abused as output transformer) is of course also included.

In order **to preempt misunderstandings**, here a short afterthought: guitar amplifiers are no HiFi-systems. The latter require a significantly wider frequency range and a significantly lower distortion. The message here is not that generally a mains transformer will do as output transformer, but rather that the high prices of output transformers result from their small production numbers, and from the more or less authentic replication of out-of-date historic examples. If authenticity is not the main objective, a mains transformer in the output stage of a guitar amplifier may be a low-cost alternative to the dedicated special output transformer. Each of us has to find out (**!**) on his/her own what is deemed suitable – the expectations vary too much.

A peculiarity: the special **JTM-45**-transformers wound – very authentically – to an R_{aa} = 8 kΩ specification. This certainly is inappropriate for an EL34-power-stage, but in the original JTM-45 we do not find EL34's but two KT-66's. Do these then require an 8-kΩ-transformer? Yes. Or no – it depends on the source. According to Doyle's Marshall-book, **Radiospares** was the first purveyor to the court with their "De Luxe Output Transformer". Radiospares, however, was not a manufacturer but a distributor (they became RS-Components later). Who actually manufactured these early transformers is the object of escalating discussions (allegedly up to five manufacturers may be in the running). The RS-transformer was a typical universal transformer featuring a choice of several primary impedances: 6.6 kΩ (with ultra-linear tap) for EL34 and KT66, and 8 kΩ or 9 kΩ for 6L6, 6V6 and EL84. The power stage of the JTM-45 does not operate in ultra-linear mode; the experts consider 2xKT-66 / R_{aa} = 8 kΩ to be the nominal complement. The Drake-transformer used after the RS-transformers operates with this primary impedance, too. And the GEC-datasheet of the KT-66 (1956), as well, specifies 8 kΩ, but does this for "cathode-bias" which is not used for the JTM-45.

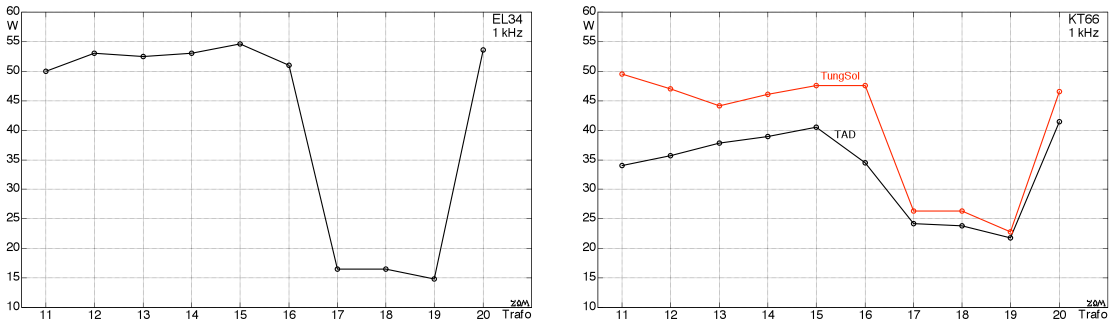


Fig. 10.6.37: Output power at 8 Ω, for a distortion-suppression of 30 dB; transformer-numbers acc. to the table.

Fig. 10.6.37 shows the measurement results for a supply-voltage of 400 V. Two EL34 will yield well over 50 W, given a primary impedance of about 3.5 kΩ. At 8 kΩ, the power output drops to a meager 15 – 16 W – for sure, this is not optimal. Using two KT-66's, about 25 W are achieved with 8 kΩ impedance, which is about in agreement with the datasheet. We obtained more power operating our JTM-45 with two KT-66's and a 3.5-kΩ-transformer: just under 50 W with a Russian TungSol-KT-66, significantly less with a TAD-KT-66 (measurement results in Fig. 10.11.3).

Besides the maximum output power, the source impedance shows differences, as well. Pentodes are of high impedance, and therefore the source impedance of the power stage (the internal impedance) is relatively high, too. It will be around 100 – 200 Ω with two KT-66 cooperating with a 3.5-kΩ-transformer, but only 40 – 80 Ω with an 8-kΩ-transformer (each at the 8-Ω-output with the negative feedback disabled). The effects have already been discussed several times; they show up e.g. in Fig. 10.6.34.

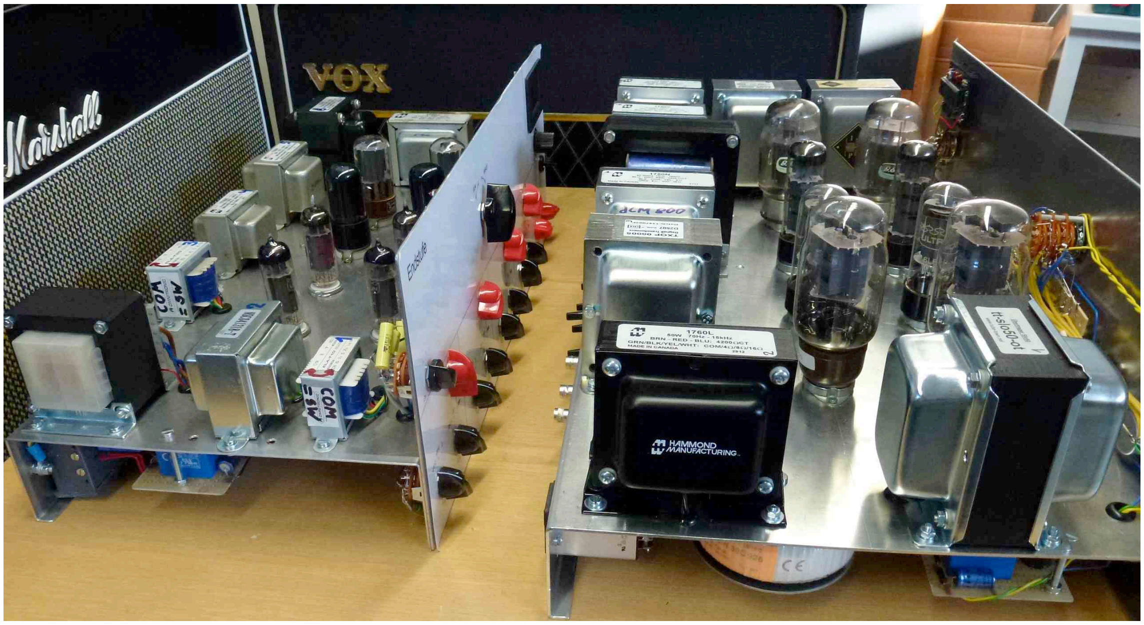
Well then, it's getting to be after hours – time to go home for dinner. It's been quite a while. You want a recommendation? Because, according to an OECD-study, many readers have difficulty to hang in there when confronted with longer texts? Ok, here we go:

Loud = 2xEL34 with 3.5-kΩ output transformer;
Authentic = 2xKT66 with 8-kΩ output transformer;
Prepared to take a risk = 2xEL34 with (special) mains transformer as output transformer;
Moronic = expensive replacement transformer from faraway lands.

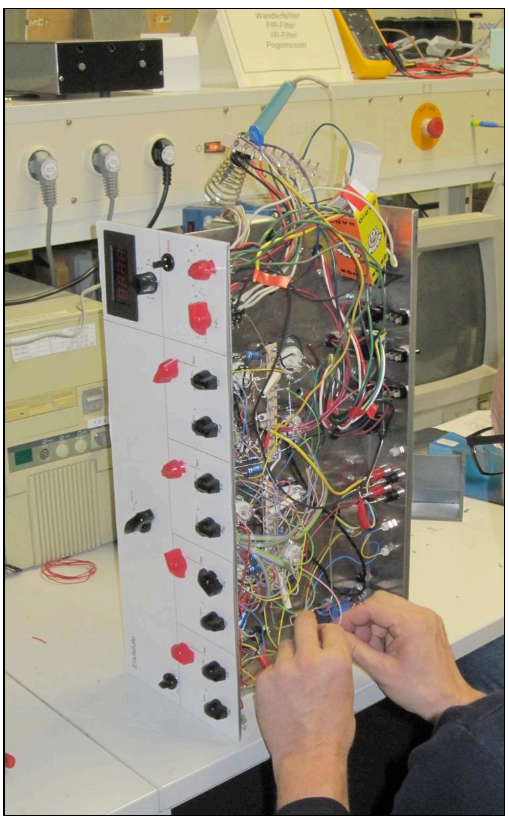
Is that short enough, and intelligible despite three multiplication signs?
You are welcome – happy to comply.

It is not the things that delight us,
but the opinion we have about the things♣

♣ loosely based on Epiktet



Power stages for the practical course on tube amplifiers



At work in the tube lab

with a somewhat disengaged participant