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).

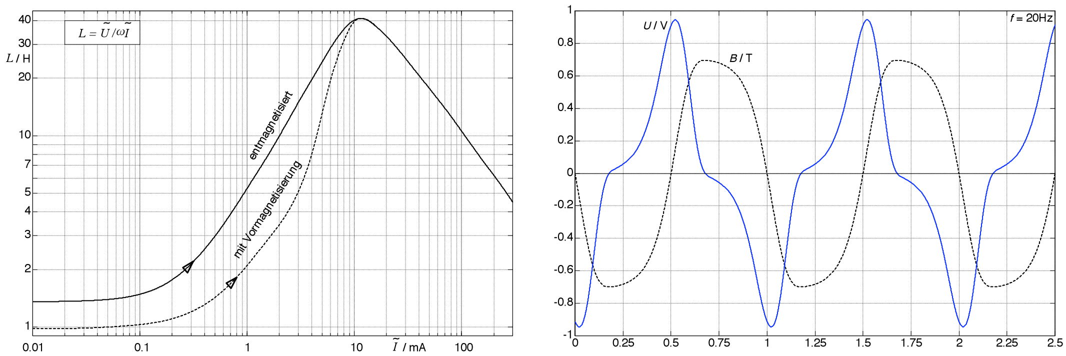


Fig. 10.6.11: Measurement at the primary winding (EI-96). The "inductance" given in the section on the left is a special non-linear quantity. Right: secondary voltage (LL) and flux-density for input from a stiff current source.

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).

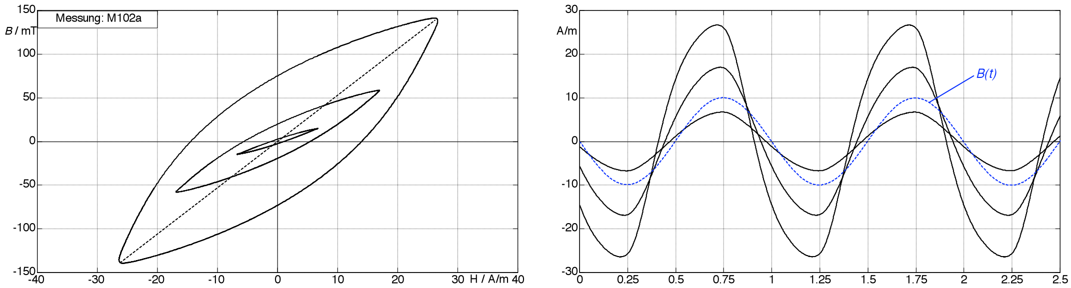


Fig. 10.6.12: Hysteresis loops. Right: time-functions of field-strength measured with imprinted sinusoidal flux-density; dashed: the time-curve of a flux-density (no scaling).

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).

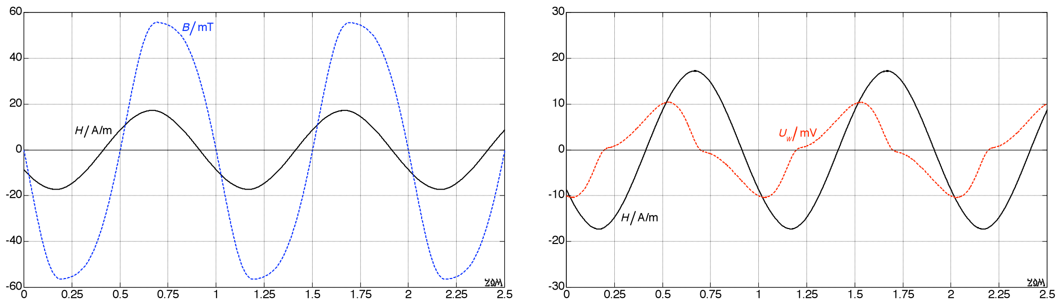


Fig. 10.6.13: Sinusoidal field-strength H (imprinted via the primary current) and corresp. flux-density B (left); non-linear distortion in the voltage U_w across the winding resulting from this H and B (right).

♣ 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 ($\approx$ 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' = $ü^2 \cdot (R_2 + R_L)$ 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.

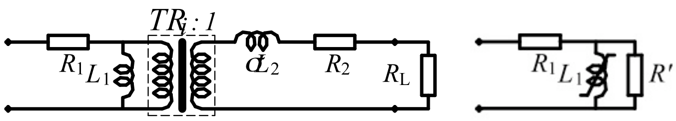


Fig. 10.6.14: Equivalent circuit for the transformer (left); two-pole simplification for low frequencies (right).

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.

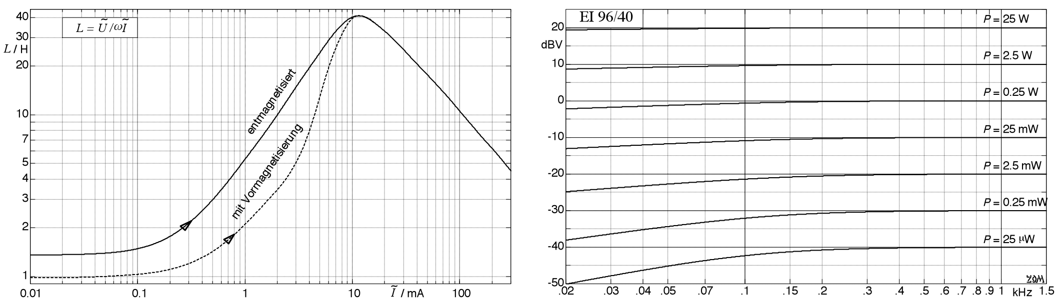


Fig. 10.6.15: Left: drive-dependent main inductance (—— core demagnetized, ----- with remanence).
Right: drive-dependent non-linear high-pass (fed from a stiff-current-source, core of transformer demagnetized). The specified power is fed to the ohmic nominal impedance (4 Ω) at 1 kHz.

♣ 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.

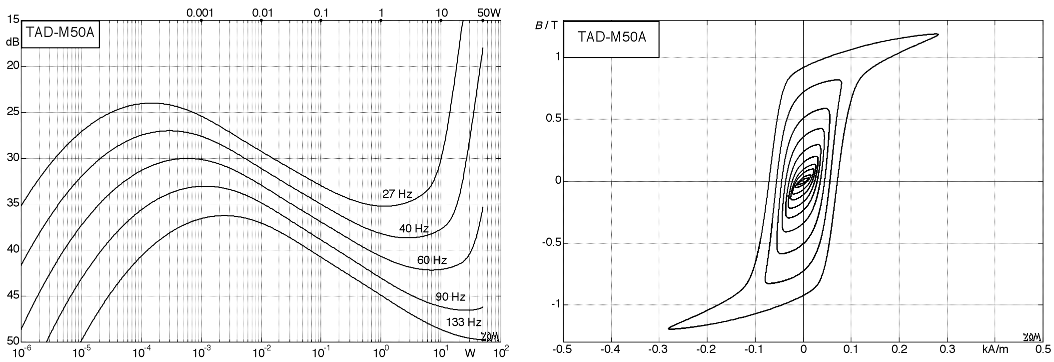


Abb. 10.6.16: Distortion-suppression a_{k3} of a 50W-output-transformer for high-impedance drive-signals and nominal load. The non-linear distortion is generated exclusively by the transformer and not by the driving amplifier. The hysteresis loop shows the relation between magnetic field-strength and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

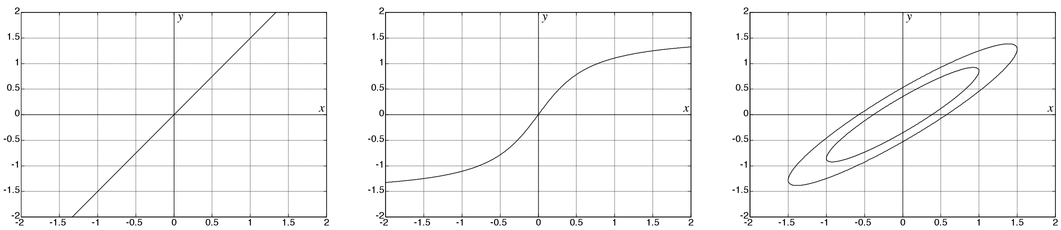


Fig. 10.6.17: Transmission characteristic of a linear system (left) and of a non-linear system (center). For dynamic (memory-containing) systems (right) two drive-levels are depicted.

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.

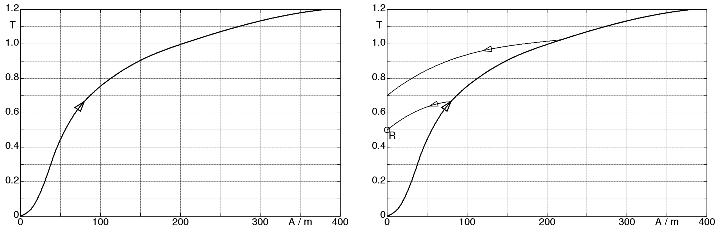


Fig. 10.6.18: Relationship between magnetic field-strength *H* and magnetic flux-density *B*.

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.

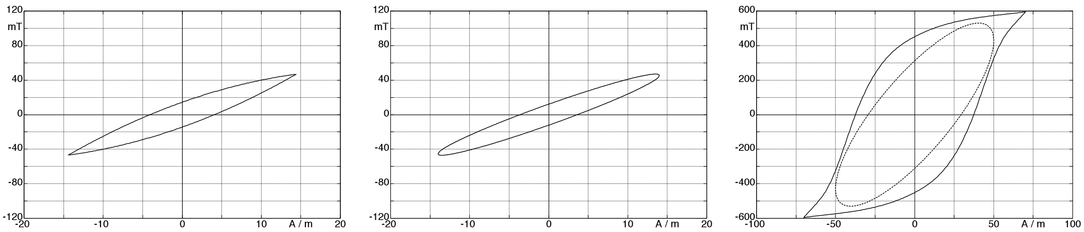


Fig. 10.6.19: Approximations using parabola (right) and ellipse (center). Limits of the ellipse-approximation as saturation sets in (right).

♣ 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.

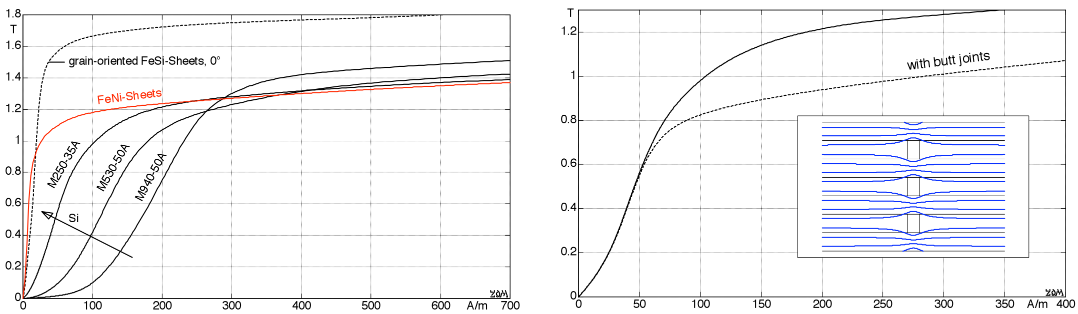


Fig. 10.6.20: Magnetic commutation curves of various core metal sheets; impact of the butt joints.

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 $\hat{B} = 1\mathrm{T}$, 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.

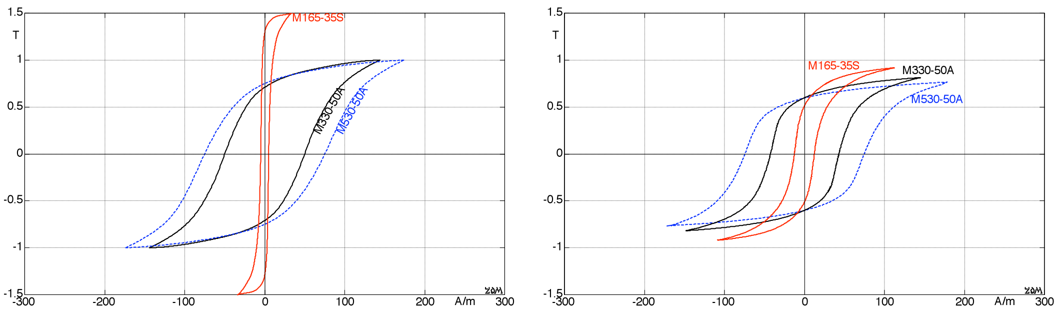


Fig. 10.6.21: Material characteristics (Waasner, left), measurements (EI96a, right). The material characteristics are valid for the base-materials; stamping will change the values; for the influence of butt joints: see Fig. 10.6.20.

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.

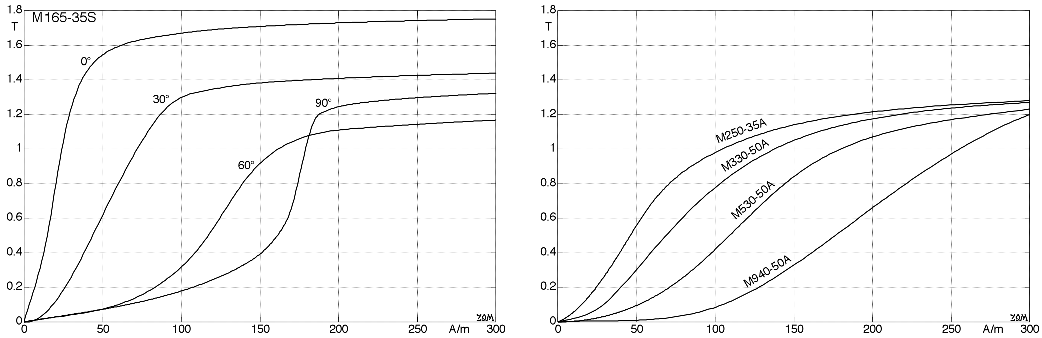


Fig. 10.6.22: Magnetization curves: grain-oriented sheets (left), isotropic sheets (right); base material.

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** (($U/\omega I$, 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.

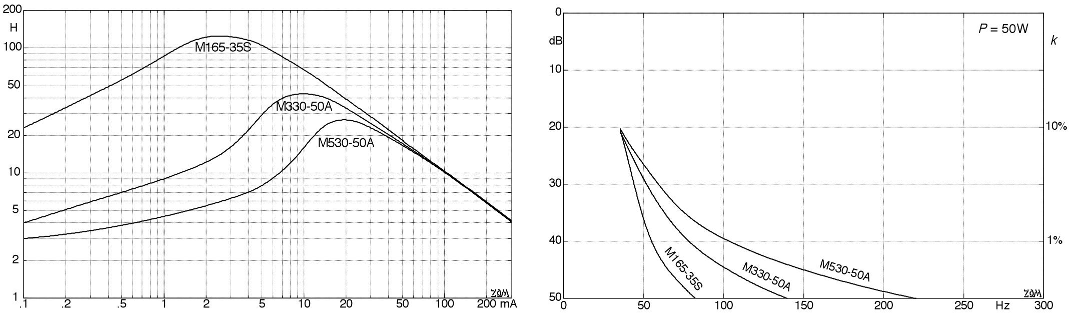
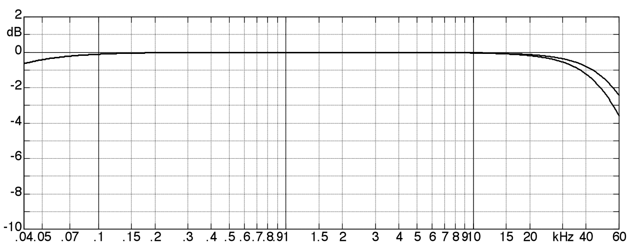


Fig. 10.6.23: Inductance of *one* primary winding (N = 900, RMS current); distortion-suppression at P = 50 W. M330-50A and M530-50A are isotropic FeSi-sheets, M165-35S is a grain-oriented FeSi-sheet. EI-96a.

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.

