10.8 Effects

10.8.1 Reverb

In every (interior) space♣, floor, ceiling and walls reflect sound. Individual reflections with a temporal distance of more than about 50 ms are perceived as individual echoes. Reflections arriving with smaller intervals create a perception of reverb. Since sound propagation in a room (i.e. in air) is – with very good approximation – a linear process (LTI-system), the system "room" may be described by its impulse response (**Fig. 10.8.1**), or by its transfer function.

In order to obtain the **impulse-response of a room**, the room is acoustically excited via a sound-impulse: this could be an electrically generated spark (spark-plug), or a bursting air-balloon, or a hand-clap, or something similar. In reality, such an excitation signal is not the Dirac impulse known from systems theory but a real sound impulse of a duration larger than zero and an amplitude smaller than infinity. A microphone picks up the sound pressure at the measuring location and its magnitude is depicted as a graph over time (**Fig. 10.8.1**). Since there are any number of excitation- and measuring-locations in a room, there is also a corresponding multitude of impulse responses. With passing time, the reflections become weaker and their density increases. The first reflections (early reflections) serve the auditory system to obtain information about the size of the room. The speed at which the reflections decrease is a measure for the absorption in the room. Slow decay results in a *reverberant* impression, the opposite impression is called *dry*.

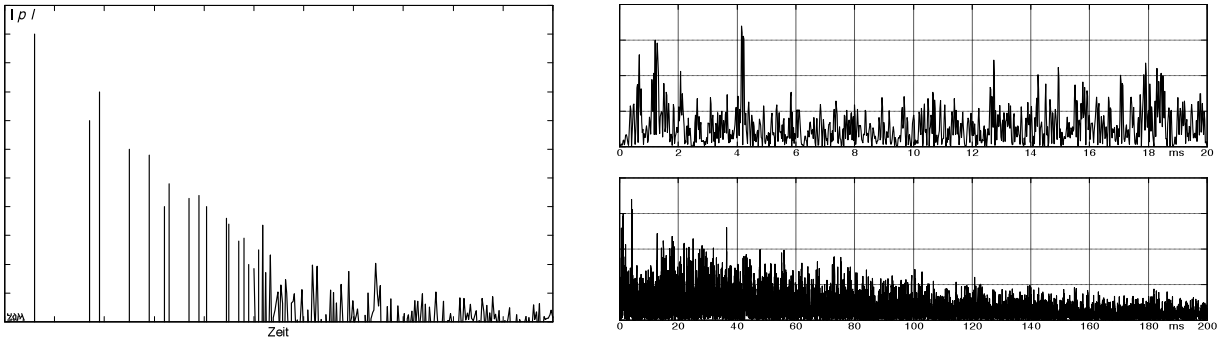

Fig. 10.8.1: Impulse response of a room (reflectogram). Left: model, right: examples for a real room.

Reverb-springs, reverb-plates, magnetic-tape devices or electronic delay-systems are used to simulate real room reverb. For guitar amps, the reverb spring has established itself as a standard from the early 1960s. After being available for some time in a stand-alone device only, it was first integrated into Fender's Vibroverb-Amp from 1963. Two steel-wires wound into helical springs serve as delay lines with mechanical waves running back and forth within them. The basics of this delay-principle had been investigated at Bell Labs, and employees at the Hammond Company had developed it into a product ready for series production. The "reverb can" (or "reverb pan" or "reverb tank") as it is manufactured today by Accutronics holds steel wires of a diameter of 0,4 mm that are wound into a helix of 4.2 mm outer diameter. On the side of the actuator, an electromagnet creates forces within a small permanent magnet that deforms the wire; the sensor side operates similarly: the movement of a permanent magnet induces a voltage in a sensor-coil.

♣ In the following, the term „room" is used; it is always associated with a space having reflecting boundaries (room, hall).

The originally used reverb system included 2 subdivided springs; newer systems are also available with 3 springs (**Fig. 10.8.2**). The connecting point (which is not located exactly in the middle) between the subdivisions creates additional reflections. As a current flows through the actuator coil, the permanent magnet creates torsion in the steel wire that results in a flexural wave. The latter runs along the wire and reaches the other end after about 30 – 40 ms. With the two springs having slightly different dimensions, the delays differ as well (43 ms and 41 ms, according to the manufacturer).

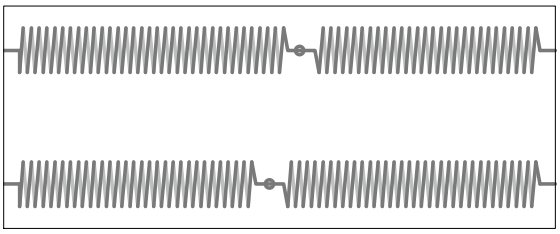

Fig. 10.8.2: Reverb system with 2 subdivided springs (Accutronics). Alternatively, systems with 3 springs are also in use. The flexural waves travel along the springs and are reflected at both ends. At the connecting points, reflections are created, as well – these are, however, less pronounced.

Compared to real room reverb, spring-reverb shows a significantly different behavior: sound propagation in space is three-dimensional and non-dispersive, while in the spring, it is one-dimensional and dispersive. If we had only a single spring without subdivision, we would get merely a sequence of echoes (e.g. after 30, 90, 150, 210 ... ms). These echoes would be equidistant, separated by the time it takes the sound to travel back and forth in the spring. Conversely, the average reflection density in a real room increases with time squared t^2. Each reverb spring consist of two parts connected via a ring. If we take the spring to be a mechanical line (compare to Chapter 2), the ring acts as mass loading which reflects in particular the higher-frequency waves. Thus, two echo-systems are connected in series, and the reflected wave obtains a t-proportional component at high frequencies. The second (subdivided) spring connected in parallel, on the other hand, merely doubles the density of the reflections without changing the exponent of the time-dependency.

Fig. 10.8.3 shows spectrograms of the impulse response: in the left picture for the two-spring system, and in the right picture for the same system but with on spring clamped down (such that no vibration could be formed). Clearly, there is not really any <u>one</u> delay-time per spring. Rather, a frequency dependent group-delay is created due to the dispersive propagation: high-frequency components require about 50% more time to arrive than low-frequency components.

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Fig. 10.8.3: DFT-spectrograms of the impulse response: two reverb-springs in parallel (left), one spring (right).
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The usable upper frequency limit of a spring-reverb is about 5 kHz – fully adequate for a guitar amp. **Fig. 10.8.4** depicts a third-octave analysis taken from an excitation with pink noise [3]. The different curve shapes in the high-frequency range result from different loading of the sensor coil: the inductive source impedance acts, in conjunction with the load capacitance, as a second-order low-pass. This low-pass can generate a slight resonance peak at 4,5 kHz (----) given the appropriate dimensioning. A measurement with a sine sweep enables us to take a closer look at the fine structure but requires consideration of the extremely long attack and decay times. Even using a sweep-duration of 2 minutes, the system cannot actually "settle": the exact position of the maxima and – in particular – the minima depends on the measurement parameters (resolution, sweep duration).

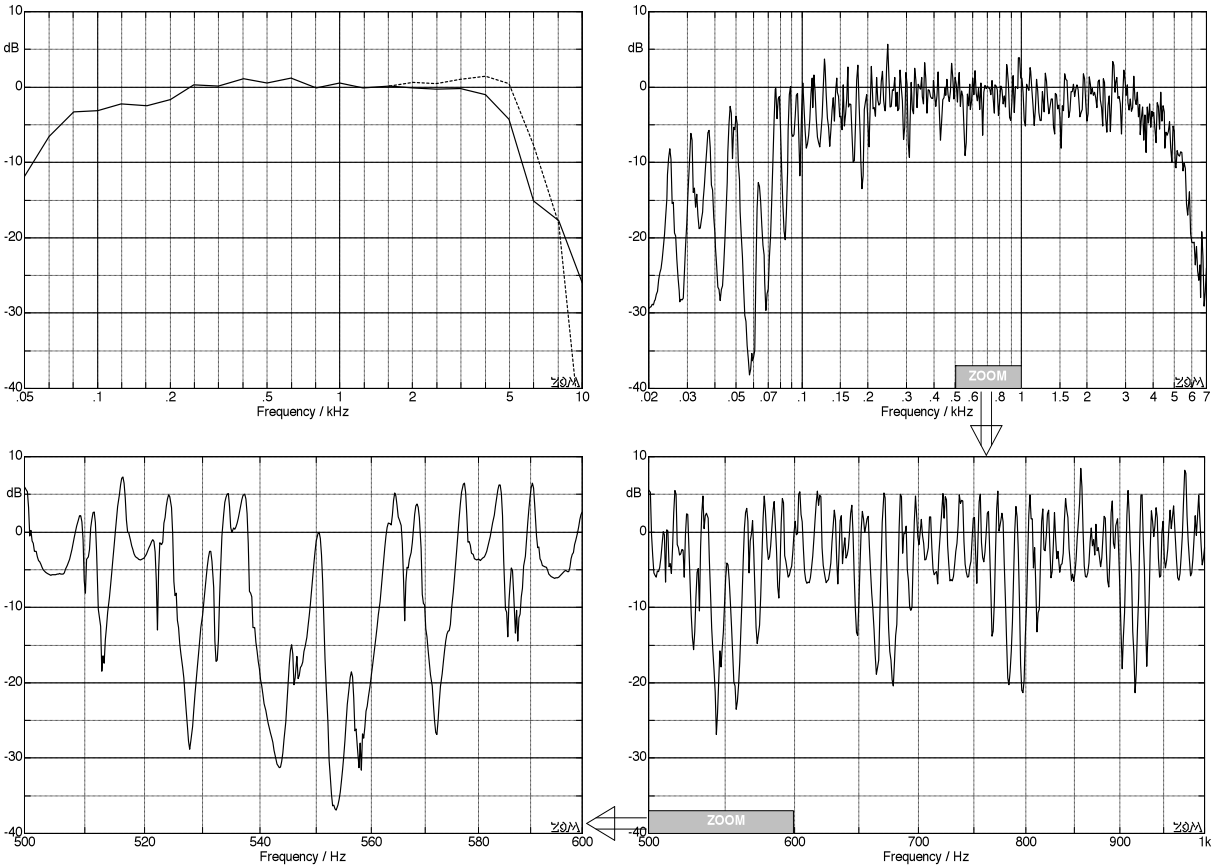

Fig. 10.8.4: Accutronics reverb 4AB3C1B, current input. 1/3-octave-analysis (upper left), sweep measurements.

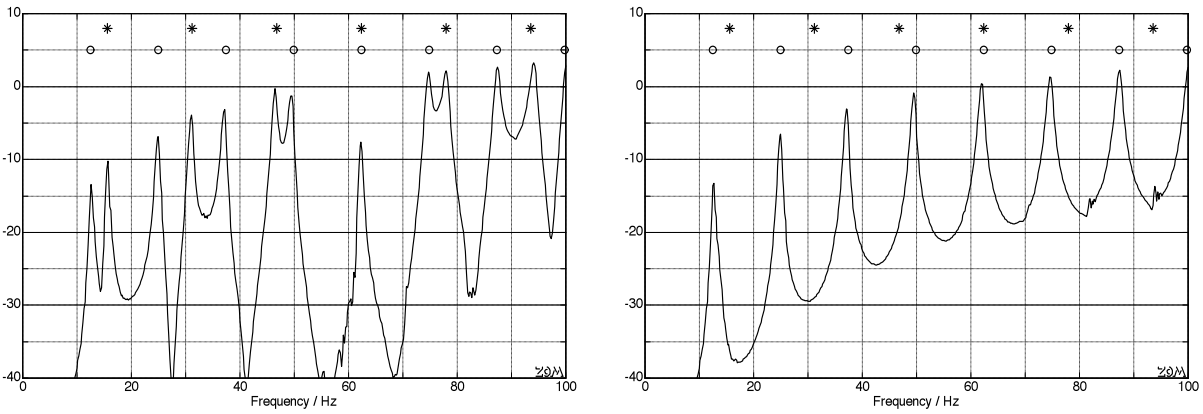

Fig. 10.8.5: 4AB3C1B, sweep-analysis; 2 springs in parallel (left), one of the 2 dampened (right).

In **Fig. 10.8.5** we see the lower-order natural frequencies with enlarged scaling. The lowest of these natural frequencies are located at 12,5 and 15,6 Hz, respectively, for the two parallel springs, and the higher natural frequencies are found at multiple integers thereof (circles and stars in the figure). At 63,5 Hz, there is an interaction of the 5th and 4th natural frequencies, the result being a beat-effect of a very long periodicity. The right hand section of the figure clarifies that the minima are the effect of destructive interference: with only one spring active, the comb-structure is much more even. This regularity also supports the hypothesis (on the basis of transmission-line-theory) that the ring positioned in the middle of the string (and connecting the string subdivisions) works as a scattering body predominantly at high frequencies.

The decay of the reverb is usually expressed as the **reverberation time** T_{60}: this is the time it takes for the (1/3-octave) level to decrease by 60 dB after switching off the excitation signal. In the high-frequency range, we find a by-the-book behavior (**Fig. 10.8.6**): the level decreases linearly with time. The reverberation time is 2,5 s. At low frequencies two superimposed decay-processes reveal themselves: an early fast decay and a subsequent slower decay. In such cases the perception-relevant **early-decay-time** is specified as six times the duration it takes the signal to decay from -5 dB to -15 dB. The reference level is the averaged level for the steady state excitation. We can see in the figure that this EDT (T_{10}) is, again, 2,5 s for the chosen 1/3-octave band – the subsequent slow decay can be attributed a reverberation time of about 12 s. In guitar amps, the frequency range below about 300 Hz does not have a particular importance (for the reverb signal): usually a high-pass will effectively dampen the lows in order to suppress any annoying booming.

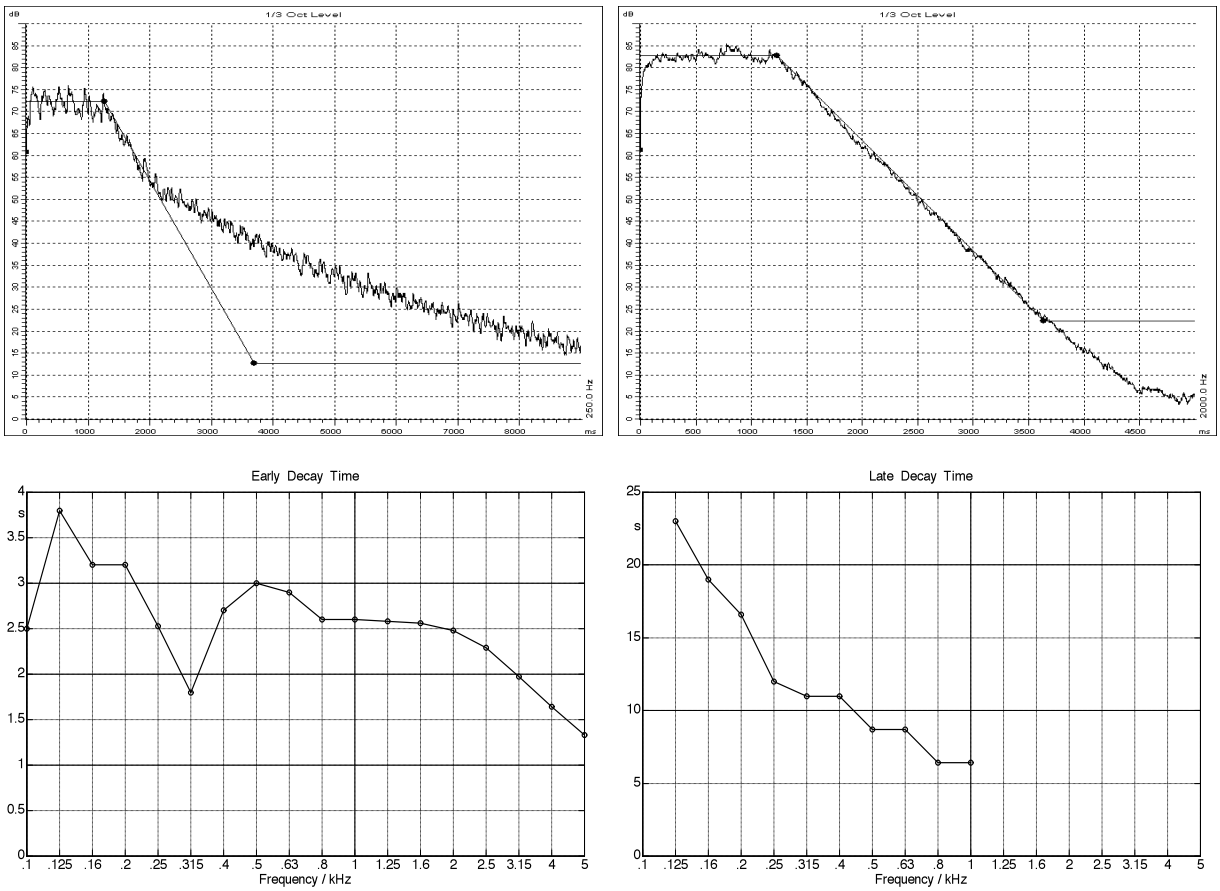

Fig. 10.8.6: 1/3-octave decay analysis, Accutronics spring-reverb 4AB3C1B. 250 Hz (left), 2000 Hz (right). The lower pictures show the frequency-dependency of the reverberation times.

The **kink** in the reverberation curve is probably an effect of the scattering mass in the middle of the spring. The waves reflected at it reach the (partially absorbing) end of the spring twice as often as the waves passing through, and we get two types of standing waves. This hypothesis, however, was not extensively tested via experiments – a full dampening of one of the two springs did in any case reveal clearly that the kink is not a result of different absorption in the two springs: the *individual* (subdivided) spring shows the same kink-behavior as shown in Fig. 10.8.6.

In order to achieve a reasonably frequency-independent transmission with the investigated Accutronics reverb system, driving it with a **stiff current source** is conducive (Fig. 10.8.4). With a voltage source, a strong treble-loss would occur due to the substantially inductive input impedance of the actuator coil. A high impedance source is automatically made available by tubes in a common-cathode circuit. However, the Accutronics system has such a low input impedance (8 Ω at 1 kHz) that an extreme mismatch would result. The optimum current drive happens with a source impedance of about 100 Ω – this is obtainable from a tube plate only using a transformer. Impedances are transformed with the transformation-ratio squared: a transformer with a 25:1 ratio will yield – for a tube output impedance of 62,5 kΩ – the appropriate secondary source impedance of 100 Ω. Fender's stand-alone reverb unit **6G-15** employs a 6K6-GT to drive the reverb-transformer; this low-power pentode features an internal impedance♣ of 90 kΩ (the 6V6-GT has 50 kΩ). If the reverb is integrated into a guitar amplifier, it is almost always a **12AT7** that is deployed; it has an internal impedance of merely about 40 kΩ per triode. Since both triodes in the tube are connected in parallel (!), the source impedance drops to 20 kΩ. On top of this, the reverb transformer 125A20B has a transformation ratio of 50. As a consequence, the reverb system is effectively driven from a voltage source above 1 kHz, and a corresponding treble-loss.

If the reverb system were a linear device (in the sense used in systems theory), we could insert a corrective filter at any point in the amplification chain and boost the missing treble. However, both the reverb spring and the tubes are **non-linear** devices, distorting at high signal levels, and generating noise and rumble in the small-signal range. Filter-design therefore always includes a component of dynamics-optimization, as well. In a typical Fender amp, predominantly the low frequencies are attenuated ahead and after the reverb spring – the treble-boost-enabling current-drive is only rudimentarily taken advantage of.

In **Fig. 10.8.7** we see the transmission frequency response from the ECC83 ahead of the reverb branch up to the plate of the ECC81. The filtering is done via two sections: the RC-high-pass ahead of the ECC81 (f_g = 320 Hz), and the inductive input impedance of the reverb-transformer. Since the double-ECC81 has quite a low output impedance, we see the reverb driven by a current source only up to about 1 kHz – in the upper frequency range, the tube acts as a voltage source. The voltage transmission-factor of the reverb system is shown in **Fig. 10.8.9** – in contrast to the situation with a current source (Fig. 10.8.4) we find a pronounced treble-loss. The connection to the reverb-potentiometer is done via a 3-nF-capacitor resulting in another high-pass filtering (350 Hz). The overall reverb-branch has a bandpass characteristic centered around 600 Hz while the direct signal only receives a mild treble-boost. The circuit depicted in **Fig. 10.8.8** is typical for many Fender amplifiers – some do have a 2-nF-capcitor connected in parallel to the reverb-pan output in order to create a small resonance peak (Fig. 10.8.4).

♣ Data-sheet specifications. ♠The 12AT7 (= ECC81) is working at an atypical operating point!

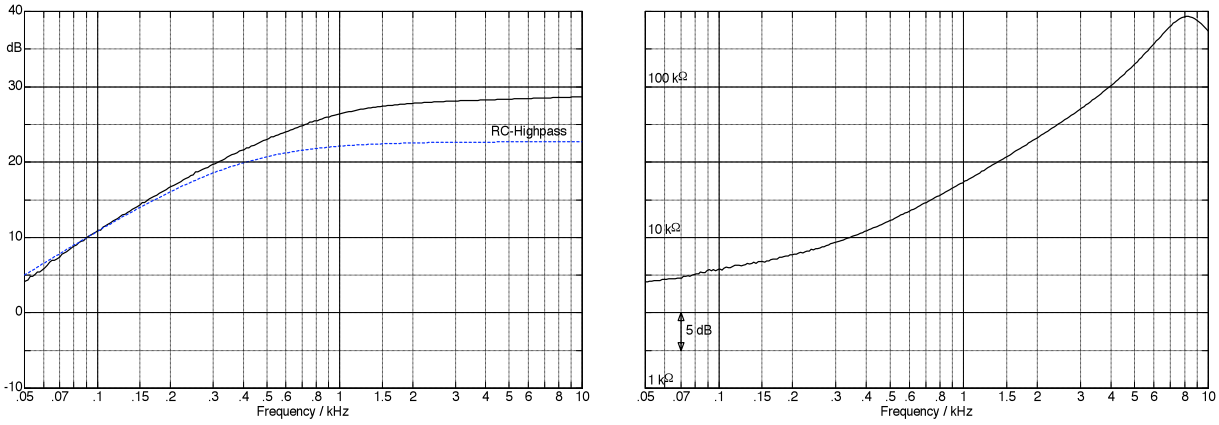

Fig. 10.8.7: Transfer characteristic from the ECC83 up to the ECC81 (left); transformer input (right).

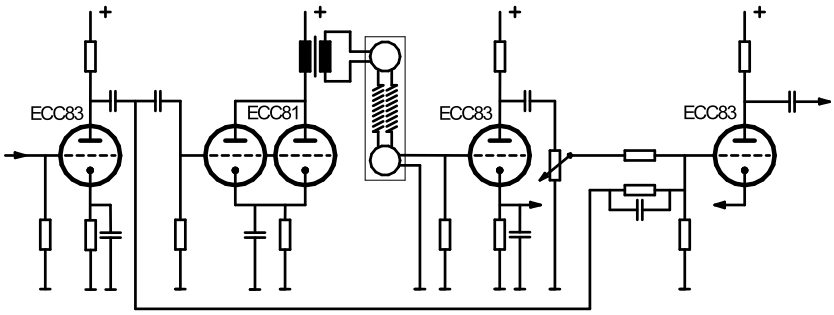

Fig. 10.8.8: Circuit of a typical Fender reverb. The reverb-spring input is of low impedance, the reverb transformer has a 50:1-ratio.

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Fig. 10.8.9: Smoothed voltage transmission factor of the spring-reverb system (left). Transmission factor of the overall reverb-branch (right ——), and of the direct signal (right -----). Both curves in the right-hand picture show the transmission from the plate of the ECC83 ahead of the reverb branch up to the last ECC83 in the reverb branch. (Fig. 10.8.8). **These figures are reserved for the print-version of this book.**

The **Accutronics reverb-pans** are coded with 7 characters, e.g. 4AB3C1B. The individual characters indicate:
1st position: type. At the time of writing the Types 1, 4, 8 and 9 are available.

Accutronics reverb-pan Type 1 and Type 4			Accutronics reverb-pan Type 8 and Type 9		
2nd position = Z_{in}		3rd position = Z_{out}	2nd position = Z_{in}		3rd position = Z_{out}
A = 8 Ω	D = 250 Ω	A = 500 Ω	A = 10 Ω	D = 310 Ω	A = 600 Ω
B = 150 Ω	E = 600 Ω	B = 2250 Ω	B = 190 Ω	E = 800 Ω	B = 2575 Ω
C = 200 Ω	F = 1475 Ω	C = 10 kΩ	C = 240 Ω	F = 1925 Ω	C = 12 kΩ

4th position = reverberation time: 1 = short, 2 = medium, 3 = long.
5th position = chassis connected to: A = Input + Output; B = Input; C = Output; D = chassis insulated.
6th position = pan lock: 1 = no lock.
7th position = preferred mounting orientation: A = ⊔; B = ⊓; C = ⊏; D = ⊐; E = IN; F = IN;

1

2

3

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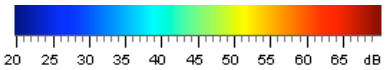


These figures are reserved for the print-version of this book.

1: Accutronics (4AB3C1B), 2 springs, subdivided.
2: Accutronics (9AB2C1B), 3 springs, subdivided.
3: low-cost, 2 springs (17 cm), not subdivided.
4: low-cost, 2 springs (14 cm), not subdivided.
5: digital reverb, "Spring-Reverb".
6: digital reverb, "Spring-Reverb".
7: digital reverb, "Room-Reverb".

Fig. 10.8.10: Various reverb spectra.

In **Fig. 10.8.10** we find a comparison between the reverb spectra of various reverb systems. The two Accutronics tanks represent the **spring-reverb** standard: they measure about 36 cm and are subdivided in the middle, with clearly visible individual reflections and dispersion. The spring-reverbs 3 and 4 are only of half the size compared to the Accutronics; they have no subdivision. The dispersion here is much stronger compared to systems 1 and 2, and the bandwidth is smaller, as is the reflection density – and the price is, of course, lower also. The pictures 5 to 7 show spectra of **digital reverb systems**. System 5 is marketed as "Spring Reverb" but has little similarity to an actual spring-reverb. In system 6, we may at least surmise that the developers sought to model the spring-typical dispersion although the result is not very authentic. System 7 is offered as "Room Reverb" and it does differ from the previously shown systems in that the strong periodicity is gone. The limited bandwidth of only 2,5 kHz is probably due to the computation power: the larger the bandwidth, the more load on the signal processor. The reverb spectra of a real room (**Fig. 10.8.11**) show, in comparison to the models, a much higher reflection-density and no discernible periodicity. The spectrogram has only limited meaningfulness here: since the DFT on which it is based cannot provide a high selectivity at the same time in both the frequency-domain and the time-domain. Nevertheless, the spectra shown enable us to get a basic insight into the individual reverb structures.

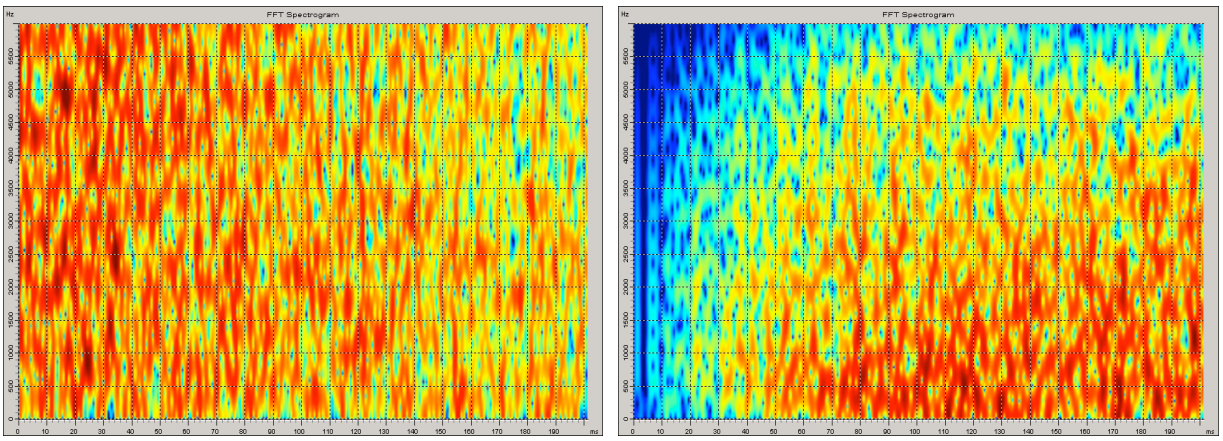


Fig. 10.8.11: Broad-band spectrum of a real room (left). Digital reverb of a studio-grade effects-processor, "large room" program with pre-delay and slight treble attenuation (left)

We should stress that every one of the reverb-systems discussed here can serve to generate a quite useful reverb for guitar. The responses following a short impulse may sound somewhat strange, but with a guitar such an excitation does not normally occur. Of course, compared to a real room, the wavering wash created by a spring-reverb has a somewhat outlandish sound at the first moment. However, the "room reverb" sounds just as peculiar in comparison if we have just listened to a Fender spring-reverb. There is a good reason that professional reverb devices offer a multitude of special reverb parameters to adjust such that the sound can be taylored to individual tastes and needs. In most cases, it is real rooms that are to be modeled (living room, hall, church, stadium), while an authentic digital simulation of a spring-reverb is not found that often. Maybe this is the reason why there are still "real" spring-reverb system and devices on the market.

10.8.2 Vibrato / Tremolo

From a systems-theory point-of-view, a vibrato- or tremolo-system is a modulator with time-variant transfer-characteristics – it changes the signal-amplitude or -frequency. **Leo Fender's** usage of these terms for his guitars and amplifiers has created a big mix-up that the music world has not really recovered from even today; a clear assignment between the terms and the respective function has not reestablished itself: does the vibrato effect in fact change the pitch – or is it the volume that varies? Fender's Stratocaster, protected since 1954 by US-Patent No. 2741146, according to the patent description holds a "*tremolo* device" to change the pitch. 50 years later, Fender brochures still use the term in the same sense. However, in a Fender service manual from 1968, the corresponding unit on a Mustang guitar is suddenly called *vibrato*, although on the same page the term *tremolo* is used for the Stratocaster and the Bronco. Similar confusion happens with the amps: *vibrato* is originally used for amplitude-modulation (change in loudness), and the *Vibro*lux amplifier indeed includes this function. How about the *Tremo*lux? Same – it's the identical effect. Does that feel complicated? Yep, without a doubt: there is a Tremolux with a tremolo-pedal♣, and also a later one with a vibrato-pedal. And, sure enough, there is a Vibrolux-version with a tremolo-pedal – and also one with a vibrato-pedal. The circuit that generates the effect, is always based on the same principle: originally it was a time-variant grid-bias that varied the amplification factor; later a light dependent resistor (LDR) illuminated by a blinking light – in any case the typical amplitude-modulator that was most often termed "vibrato" in the Fender brochures. Not always, though: what does the 1968 Fender brochure designate the built-in amplitude modulator for the Princeton Reverb (sporting a "Vibrato" pedal)? Right you are: it is called a tremolo.

Fender did offer not just this one modulation effect: in 1959, the **Vibrasonic** amp received a special circuit generating a mixture of frequency modulation (FM) and amplitude modulation (AM). In the mid-frequency range there was mainly FM, and in the treble and bass ranges an AM working in opposite directions: as the treble got louder, the bass got softer, and vice versa. This same circuit could be found at the beginning of the 1960s also in the Concert, Bandmaster, Pro, and Super amps, and in a slightly modified version in the Showman and the Twin. Its reign was short, however: it soon was replaced by the LDR-amplitude-modulator. With one exception all these effects were designated "vibrato" at Fender; just for the Princeton reverb the same effect was called "tremolo" – as mentioned above.

In summary: at Fender, "tremolo" is often (but not always) used for FM, and "vibrato" often (but not always) stands for AM. The classical (and scientific) definition is the other way 'round: tremolo = AM, and vibrato = FM.

How are these two effects **perceived** differently in our **hearing**? Surprisingly: not to a big degree – as long as the modulation is not too strong. The reason is that pure FM does not occur in normal situations: due to selective resonances in speakers and, especially, in the rooms we listen in, FM always generates an additional AM [e.g. 3]. The latter may even be detected (for small modulation indices) more easily by the hearing system. It is difficult to generate FM with a strong modulation index while it is much easier for AM. Here we may find the reason why Fender says good-bye to FM in the early 60s, and fits the low-cost and highly efficient LDR-modulator into all his amps. The following circuit descriptions focus predominantly on the well-documented and trend-setting Fender amps – well aware that other manufacturers have also developed and successfully marketed vibrato/tremolo-circuits.

♣ Label of the footswitch-jack

In your typical tube amp, a triode generates the low-frequency signal (**LFO** = low frequency oscillator) while the modulation itself happens in another tube (or more tubes), or in the **LDR**. The LFO is of a relatively simple build: a tube in common-cathode configuration with frequency-dependent feedback. The tube inverts the signal from plate to cathode, and consequently the feedback circuit needs also to invert; both inversions result in a phase shift of 2π, which is the requirement for self-excitation. In addition, the loop gain needs to be larger than one – easily achievable with a tube. **Fig. 10.8.12** shows a circuit as it is frequently utilized (Fender, VOX, and many others). The feedback branch consists of a 3rd-order high-pass with a variable resistor that adjusts the oscillation frequency between about 3 Hz and 11 Hz. Since there is no amplitude control, the generated signal is not of perfect sine-shape – the system is non-linear and therefore there is, strictly speaking, not really a transfer function as such. This should not be seen as a problem, however, since the approximation achievable with the linear model is perfectly practice-oriented and therefore adequate for the present context.

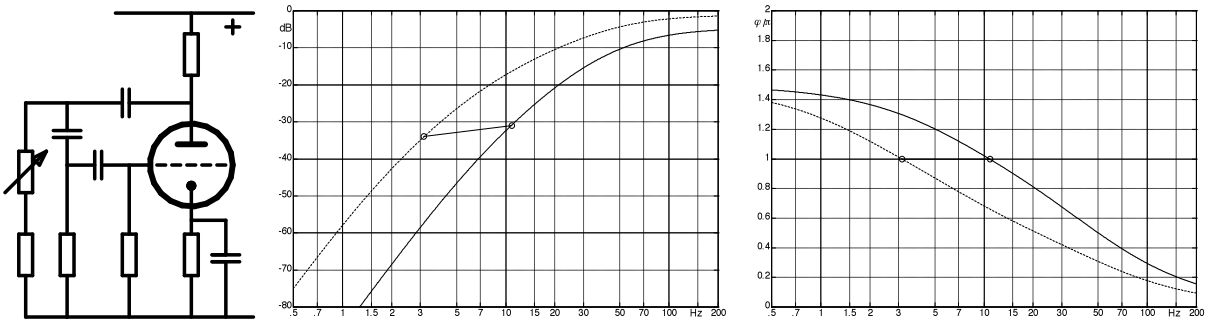


Fig. 10.8.12: LFO-circuit in a tube amplifier; magnitude- and phase- characteristics of the feedback network.

From about 1963, Fender amplifiers were fitted with an amplitude-modulator that used an opto-coupler: an **LDR** intermittently illuminated by a **glow-lamp**. The required control signal was tapped (with high impedance) from the circuit described above, and fed to the glow-lamp via the second half of the double triode (ECC83). Due to the operating point chosen for this second triode, a significant current is flowing only during a relatively short part of the LFO-period, and the glow-lamp lights up only for a short time. The resistance of the LDR decreases when lit and causes – integrated into the parallel branch of a voltage divider – a signal-attenuation (**Fig. 10.8.13**). Significant slurring of the envelope occurs due to the relatively long recovery time of the LDR – this is, however, rather beneficial to the auditory perception.

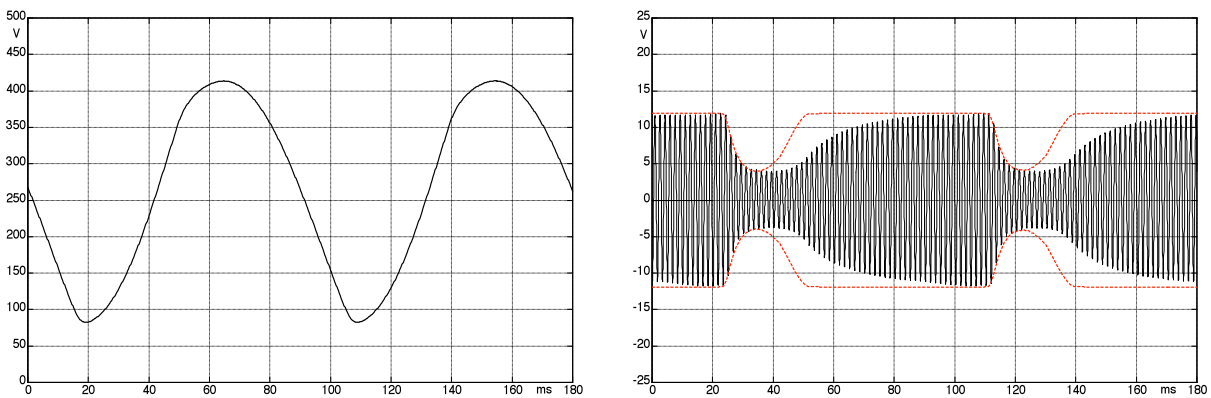


Fig. 10.8.13: LDR-modulator. LFO-signal at the plate of the oscillator-tube (left), 600-Hz-sine-tine modulated by the LDR (right). Dashed: imaginary effect of a modulator with zero recovery time.

Preceding the LDR-era, Fender deployed tube modulators. There are three types of circuits: screen-grid modulation in the power-stage, modulation in the phase-inverter, and, for single-ended power amps, modulation in the intermediate amplifier stage. The amplitude-modulation is achieved simply by shifting the operating point: a superposition (addition) of an AC voltage of very low frequency periodically moves the operating point into the end ranges of the characteristic curve. Here, the slope of the latter (and thus also the gain) is smaller than in the middle range: the gain becomes time-variant. The non-linear signal distortion also created could be accepted as an additional effect; the low-frequency parasitic signal also occurring (even without input signal from the guitar), however, requires including additional high-pass filters. For push-pull output stages, there is an elegant workaround: since the output transformer constitutes the difference of the two anti-phase signals, all common-mode signals cancel each other out (as it happens in every differential amplifier). The guitar signal is fed out-of-phase into the two halves of the push-pull stage while the LFO-signal is fed in-phase to the two sections. The result is that the guitar signal is doubled while the spurious LFO-signal is cancelled.

The **control-grid voltages**[♣] of the power tubes offer themselves as the "last possibility" to achieve the mentioned shifting of the operating point; it is implemented e.g. in the Tremolux 5G9). Synchronously pushing both grid voltages into the negative makes both tubes block: the audio signal is attenuated. Apparently, this power-tube control was seen as superior. It is found in several Fender amplifiers, and it superseded the **driver-stage control** (e.g. Tremolux 5E9-A) introduced a few years before and feeding the LFO-signal to the cathode of the phase-inverter. In both circuits an in-phase excitation of a differential amplifier is accomplished which (ideally) will avoid any LFO-signal coming out of the loudspeaker.

In the Fender Vibro-Champ (AA764), this LFO-compensation does not work because it has a single-ended power amp. Here, the LFO-signal is fed to the **cathode of the driver-tube**, and it is amplified together with the guitar signal, resulting in a low-frequency interference. The high-pass inserted directly ahead of the power-tube provides merely limited relief.

In contrast to the amplitude-modulator described above, the AM/FM-circuit first included in 1959 into the **Vibrasonic** is not understood prima facie. Here, the guitar signal is fed to a **frequency crossover** and separated into a high-pass branch and a low-pass branch[♥]. The effect is mainly a change in the loudness of the partials, but to a small degree there is also a change in phase, and therefore in pitch. The momentary angular frequency is, in fact, the derivative of the phase angle φ [3]. Since a 1st-order high-pass changes the phase by up to 90°, and a 1st-order low-pass does this by up to -90°, phase-shifts occur – as we change from the high-pass filtering to low-pass filtering – of up to about 120° (in the Fender-typical circuit). A pitch modulation with a frequency-swing of about ±10 Hz is possible with this approach, allowing for definitely audible changes in pitch. The threshold for just noticeable frequency changes is about ±2 Hz for FM-tones [12]. In **Fig. 10.8.14**, we see the magnitude and phase characteristics of the Vibrasonic-circuit (5G13); the schematic is given in **Fig. 10.8.15**. In later amplifiers (e.g. 6G13-A), the resistive voltage divider in the high-pass branch was dropped, with a gain of 7 dB in this branch.

[♣] In principle, the screen-grid voltages of the power-tubes could be modulated, as well, but this would require a higher control-power.

[♥] Strictly speaking: high-pass and bandpass, but the bandpass center frequency is, at 60 Hz, very low.

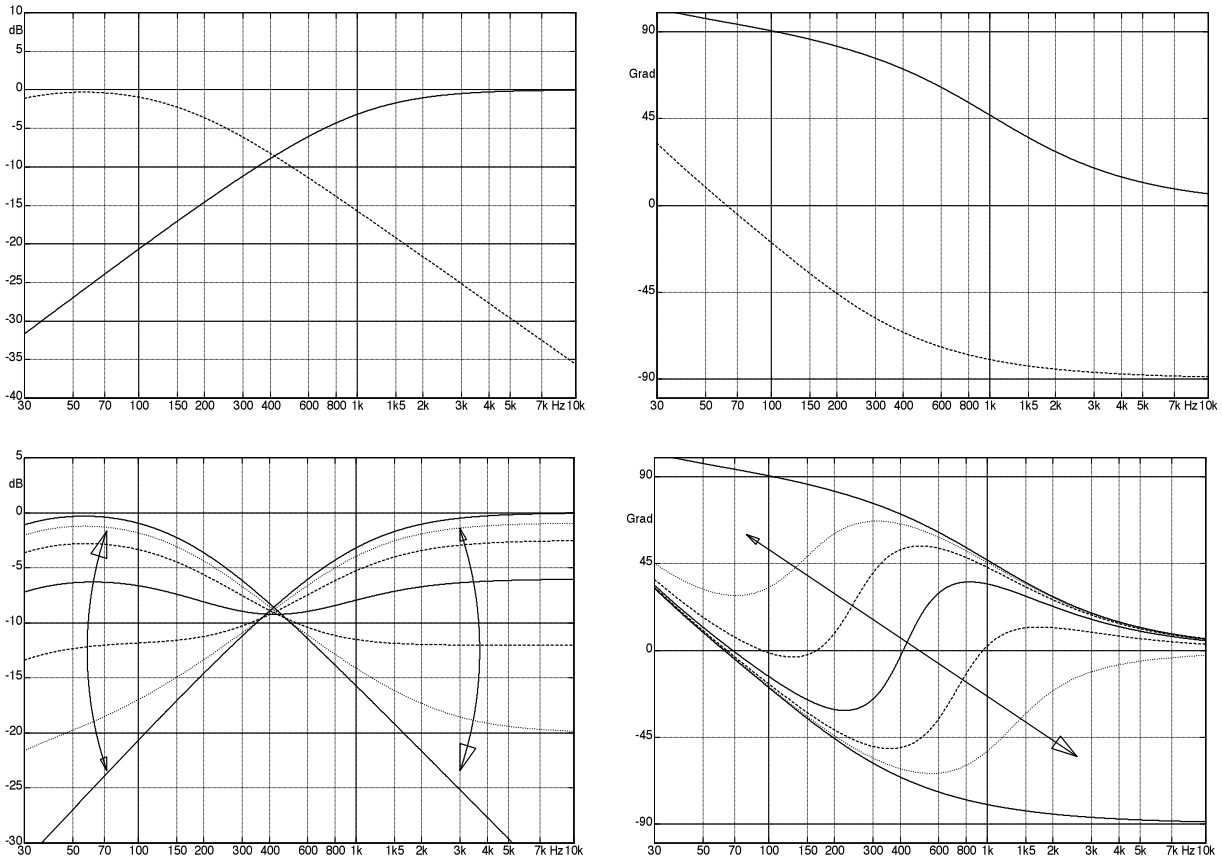


Fig. 10.8.14: Magnitude- and phase-characteristics of the frequency crossover (top), and of the overall system (bottom). 5G13.

This small **frequency modulation** realized in the Vibrasonic et al. was perfected in the **VOX** AC-30 (**Fig. 10.8.15**) using an **all-pass** circuit from the Wurlitzer organ [Petersen/Denney] that generates mainly FM but almost no AM. The required filter network is considerable: it uses 6 capacitors, 6 resistors and 3 amplifiers. It may nevertheless be divided into simple partial systems for a calculation purposes. The schematic shows two active 2nd-order bandpass filters of the same structure, differing merely in the values of the components. The signal mapping from U_0 to U_a is easily understood by omitting R_3 and C_3, to start with. What remains is a capacitively bridged voltage-divider determined by 4 components (4 degrees of freedom). One of the latter is the impedance level which tube-typically is chosen to be in the 100-kΩ-range. The second degree of freedom is the attenuation factor (about 3). Pole/zero-compensation yields the third degree of freedom ($R_1C_1 = R_2C_2$), and the cutoff frequency (about 1 kHz) yields the fourth. The result is a passive system of zero (!) order that generates a frequency-independent attenuation (of about 10 dB) across the whole frequency range.

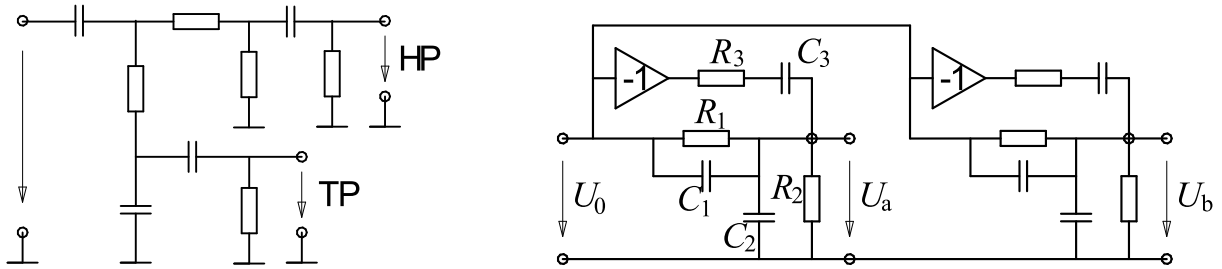


Fig. 10.8.15: Frequency crossover of the Vibrasonic 5G13 (Fender, left), and of the AC-30 (VOX, right).

Now we add R_3 and C_3, driven by an inverter ($v = -1$). Given the correct dimensioning, the output signal (U_a) is of the same phase at low and high frequencies. The active branch (R_3, C_3) cannot have an effect at low frequencies since C_3 is of high impedance relative to the divider-resistors. At high frequencies, no effect is there, either, because here R_3 is of high impedance relative to the divider-capacitors. It is only in the range of the cutoff frequency that R_3 and C_3 determine the transmission and have the effect of a phase shift from 0 to π. The components of the active branch (R_3 and C_3) can be determined such that the magnitude of the transfer characteristic becomes frequency-independent: this corresponds to a true all-pass.

An **all-pass** is a filter that changes only the signal phase but not the signal amplitude. This is achieved if the numerator of the transfer function is the complex-conjugate of the denominator of the transfer function; the magnitudes of numerator and denominator are equal for this condition, the magnitude of the transfer function becomes a constant (i.e. it is not dependent on ω). In the case of the filter circuit described above (Fig. 10.8.15), we get a 2nd-order all-pass the numerator- and denominator-polynomial of which contains p at the most with the power of two.

$$\underline{H}(p) = \frac{ap^2 + bp + 1}{cp^2 + dp + 1} \cdot H_0 \qquad a = c, \;\; b = -d. \qquad \text{Second-order all-pass, } p = j\omega$$

A 2nd-order transfer-function has 5 degrees of freedom. Two of these are required by the all-pass characteristic: the same behavior for $f \to 0$ and $f \to \infty$ results in $a = c$, and the complex conjugation of numerator and denominator yields $b = -d$. The remaining 3 degrees of freedom are defined by: basic gain (H_0), cutoff frequency (a) and Q-factor (b). The components of the AC-30-filter in the original circuit were chosen such that not a perfect all-pass resulted but a slight magnitude change did also occur (about 3 dB). The reason for this is unknown; possibly the additionally generated AM was desirable.

An all-pass in itself does, however, still not generate a frequency modulation (FM) – it only creates a stationary (time-invariant) phase shift. For this reason there is a second all-pass (Fig. 10.8.15) with a cutoff frequency of a factor of 4,5 lower than the cutoff frequency of the first all-pass (1040 Hz vs. 4700 Hz). There will be a significant phase difference between the output signals of these two all-passes that can be turned into a time-variant phase-shift by a LFO-controlled cross-fading between the two outputs. If we take the phase modulation to be approximately sine-shaped, the maximum of the frequency modulation generated this way corresponds to the product of modulation-frequency (LFO-frequency) and phase-change amplitude: $\Delta f = f_{\text{mod}} \cdot \hat{\varphi}$. With f_{mod} = 10 Hz and a maximum phase-change amplitude of 55° (= 0.3π), we obtain a frequency-change amplitude of Δf = ±9.4 Hz.

We know from psycho-acoustical experiments that the threshold for just noticeable frequency differences is about ±2 Hz at low frequencies; the **auditory system** becomes increasingly less sensitive to absolute frequency changes only at frequencies above 500 Hz [12]. The frequency modulation generated by the AC-30-Modulator is therefore clearly audible; in addition we need to consider that the modulator circuit, and loudspeaker- as well as room-resonances, additionally generate amplitude modulation. In conclusion, it should be noted that in the AC-30, one of the two all-passes can switched-off such that the amplitude modulation becomes the dominating effect.

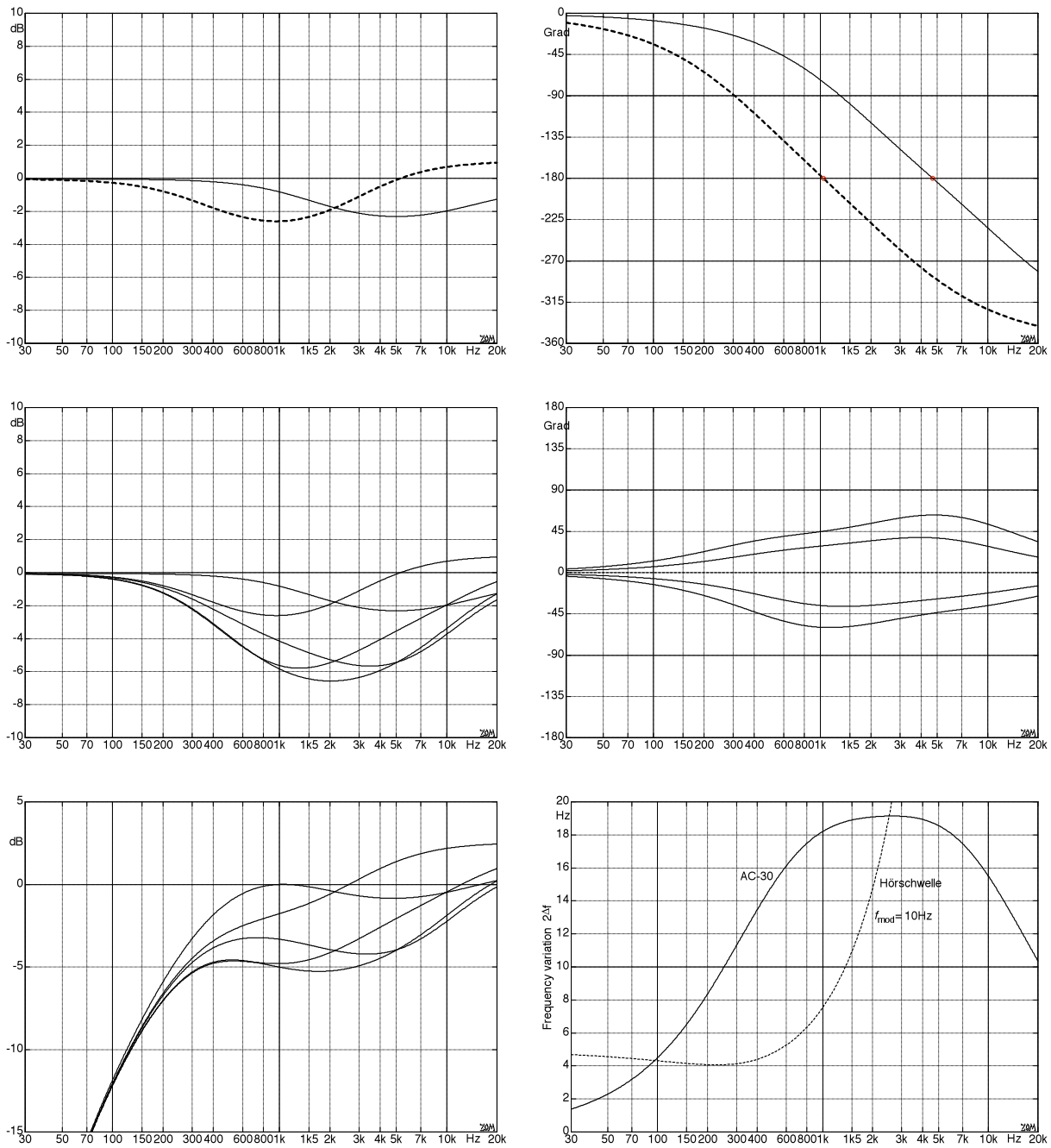


Fig. 10.8.16: AC-30-modulator. Top: magnitude and phase characteristic of both all-pass filters.
Middle: Magnitude- and normalized phase characteristic of the overall modulator.
Bottom: Magnitude characteristic incl. 4th-order high-pass; frequency change $2 \cdot \Delta f$ achievable with f_{LFO} = 10Hz.
"Hörschwelle f_{mod}=10 Hz": threshold for just noticeable FM at a modulation frequency of 10 Hz

In **Fig. 10.8.16**, calculations regarding the transfer behavior are depicted. The LFO-controlled crossover between the all-passes generates level changes of up to 5 dB, and phase changes of up to 110°, resulting in frequency changes of up to 19 Hz at 10 Hz modulation frequency. In the lower right picture, the FM-perception-threshold is shown (dashed) for comparison [12]; the achieved modulation is clearly above threshold. The 4th-order high-pass added in for the picture on the lower left follows the modulator in the AC-30 to detach the remaining LFO-signal from the guitar signal. A compensation of the LFO-signal is implemented in the summation stage but this can never be perfect due to unavoidable tube tolerances.

10.8.3 Phaser / Flanger / Chorus

Since the first electric guitars came into service, there was also the wish for sound-modifications. First, there was only a "tone"-control (a potentiometer with a capacitor), then more sophisticated sound filters were added, followed by electronic vibrato, tremolo, echo, and reverb. The typical guitar-**echo** results from periodic signal repetitions (about 50 – 500 ms delay-time), simple **reverb** combines several echo sequences of differing periodicity, high-quality reverb is generated by springs (10.8.1) or digital signal processors (in the studio, reverb-plates or reverb-chambers are used, as well). **Phaser**, **flanger** and **chorus** are electronic effects based on a short delay. The **delay** is a linear system that delays signals. A **short time-delay** sets the signal back by a few milliseconds, and therefore is different from an echo-system.

For phaser-, flanger- and chorus-devices, the delayed signal is added to the original signal such that a **comb-filter** results. The name is derived from the fact that the magnitude-frequency-response has a remote similarity to the teeth of a comb (**Fig. 10.8.17**). Plotted against a linear division of the frequency axis, the maxima and minima alternate in equal frequency distances; the figure, however, shows the logarithmic frequency scaling as it is preferred in electro-acoustics. Apart from the basic gain (not that important), two parameters determine the filter behavior.: the **delay-time** τ and the **delay gain** k. Varying τ will change the frequencies at which the maxima and minima occur (i.e. the distance between the notches in the frequency spectrum), while k governs by how many dB the gain factor changes (i.e. how deep the notches are). For a negative k, the first minimum is at $f = 0$.

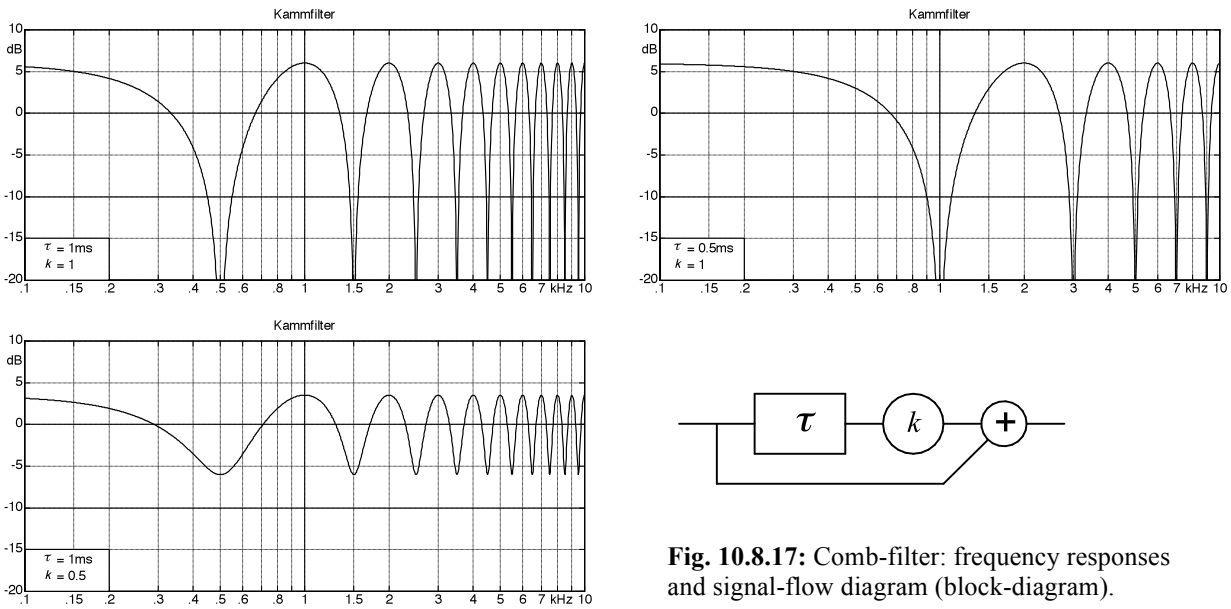


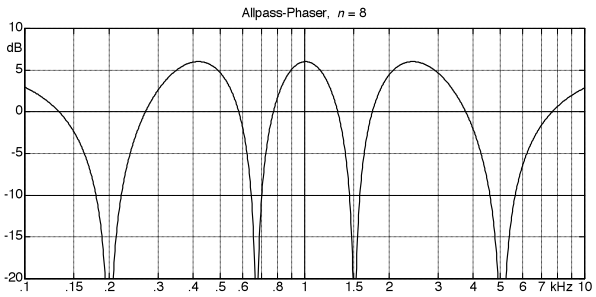
Fig. 10.8.17: Comb-filter: frequency responses and signal-flow diagram (block-diagram).

The comb-filter is in fact a typical **interference filter**: for a sine signal, a delay of half a period leads to a cancellation (or attenuation). A delay of a whole period causes amplification. Maxima and minima repeat with the period of the frequency: a minimum occurs with a delay of 1.5, 2.5, 3.5, … (or generally $n + 0.5$ with $n = 0, 1, 2, ...$) periods of the sine-signal. For the maxima, the situation is similar. In systems theory, such a filter is also termed FIR-filter, due to its impulse response which is finite in the time domain: **F**inite **I**mpulse **R**esponse filter.

The special aspect about phaser / flanger / chorus is, however, not really to be seen in the periodical frequency response but in the variation of the latter over time. A low-frequency-oscillator (**LFO**) changes the delay-time τ periodically. For example, τ swings back and forth once per second between 1 ms and 2 ms, making each minimum and each maximum sweep across a certain frequency-range as a function of time. Strictly speaking, we encounter here a time-variant system the description of which is not entirely trivial – but the quasi-stationary approximation of the shifting comb-teeth (or notches) is good enough in practice. Adding original and delayed signal (positive k) positions the lowest-frequency minimum at the inverse of twice the delay-time (1ms ⇒ 500Hz). For too short a delay-time, there is barely any audible effect because small changes occur in the high frequency region only. For a **flanger**, a typical delay-time range is 1 … 5 ms, in extreme cases this may extend from 0,3 to 15 ms.

As the delay-time is increased to above about 20 ms, a new auditory perception is generated: the **chorus-effect**. As a first-order approximation, both flanger and chorus can be described with the block-schematic as given above. Due to the very short delay-time, the flanger generates relatively broad minima in the signal spectrum and thus predominantly changes the color of the sound. Conversely, the delay-time of the chorus approaches already the value where single echoes might be discernible. This occurs at about 50 ms delay-time; our auditory system can not yet distinguish echoes as such at $\tau = 25$ ms, but it recognizes already a "fellow player". This effect is the aim of the chorus: the slightly delayed repetition is intended to create the fuller sound of not just one but two instruments playing. In addition, the delay-time is modulated by the LFO (as it is in the flanger), creating an impression of a whole instrument-ensemble. The term chorus is derived from "choir"; in the latter the individual voices start at slightly different times and sing slightly different pitches. The **pitch change** (more exactly the frequency change) is the result of the time-variant delay-time $\tau(t)$. As τ increases, f decreases, as τ decreases, f increases. The relative de-tuning is calculated as the change of the delay-time over time: $\Delta f / f = -d\tau(t)/dt$. As an example: if τ rises linearly by 10 ms within 0,5 s, the frequency of the delayed signal is decreased by 2%. A delay modulation in the shape of a triangle generates a back-and-forth sweep in the pitch. With a subtle mixing-in of the chorus (slow modulation, small frequency shift) the desired wavering choir effect is generated. For extreme settings a whining frequency modulation becomes audible.

The **phaser** is similar to the flanger but uses all-pass circuits to generate the delay; these all-passes were originally created using active circuitry (**Fig. 10.8.18**). The RC-combination determines the delay – with the R being the controllable element (as LDR or FET). Since a 1st-order all-pass can only shift the phase by 180°, several all-pass circuits need to be connected in series: n = 6…10 would be a typical number. In contrast to the flanger, the minima are not equidistant, and fewer interference notches of greater width are created..



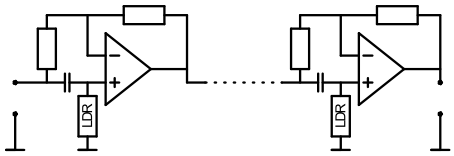
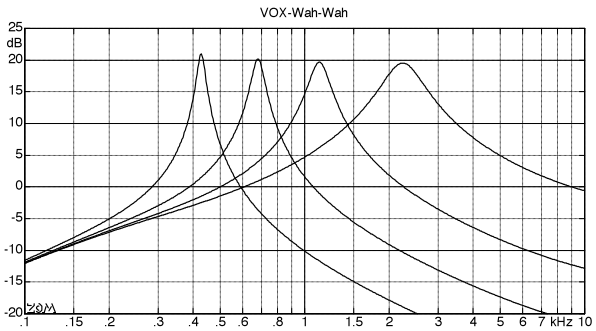


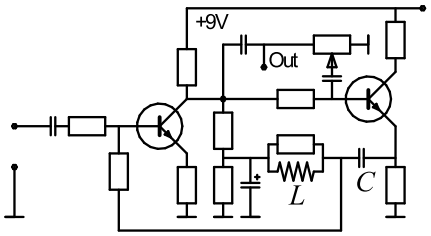
Fig. 10.8.18: All-pass phaser

10.8.4 Wah-Wah-Pedal

The Wah-Wah-pedal is an effects device performing speech-like formant-filtering. **Formants** are maxima in the speech spectrum that classify the speech-sounds [3]. The frequency F_1 of the lowest formant is about 400 Hz for the spoken vowel /o/, while for an /a/ it is around 800 Hz. If, while playing a guitar, a band-filtering is introduced with a time-variant center-frequency, and if the latter sweeps between 400 Hz and 800 Hz, we obtain a sound change that can be described by the vowel-sequence /oaoaoa/, or with “wahwahwah”.

In some devices the filtering was achieved via an LC-filter, the coil inductance of which (or rather the air gap) was variable via moving a pedal. Most Wah-pedals, however, put to use an active filter circuit in which the filter capacitance is varied by changing the gain (Miller effect). This arrangement allowed for a sweep between about 400 Hz and 2 kHz – measurements with an old VOX-wah yielded 0.44 – 2.3 kHz. The boost of the frequency range around 2 kHz is typical for the formant of an /i/, so that using the full range of the pedal results in a vowel-sequence akin to /oaiaoaiao/. More sophisticated devices (marketed with the designation 'Yoy-Yoy' or 'Doing-Doing') offered two synchronously tunable filters – presumable to more precisely imitate the human voice. Tempi passati – long bygone times.





Fi. 10.8.19: Wah-Wah-pedal (VOX).

Fig. 10.8.19 shows the circuit as well as some transmission-frequency-responses of a VOX-Wah-Wah. The inductance (about 0.5 H) and the capacitance C (10 nF) determine the centre-frequency of the filter – the capacitance is however enlarged in its effect by the factor of the gain (0…27). The capacitance effective for the filter is thus 10…280 nF resulting in a pole-frequency of 0,44…2,3 kHz. Apart from some copper- and ferrite-losses, the resistor (33 kΩ) that is connected in parallel to the coil determines the **Q-factor of the filter**; the latter also depends on the centre frequency. From the systems-theory point-of-view, a pole-Q-factor and a (different) zero-Q-factor could be specified, but in practice the “Q-factor” usually is determined using the 3-dB-down-bandwidth. For the above circuit, this definition yields Q = 3.3…15.

"**Auto-Wah**" is the designation for a Wah-Wah-filter that automatically controls its center frequency. The control parameter is the signal strength i.e. approximately the loudness of the guitar signal. Without any signal, the system tunes to the lowest possible center frequency. As the strings are plucked lightly, the centre frequency rises slightly, for strong picking the band-filter quickly sweeps from low to high frequencies and returns more slowly to the starting state. This picking-strength-dependent filter-control enables the guitar player to use the wah-wah-effect without having to operate a pedal. There will be less versatility but also less stress for the foot.

10.8.5 Distortion devices

In communication engineering, two types of distortion are specified: linear and non-linear distortions. Linear distortion is created in systems with a frequency-dependent transfer-characteristic, i.e. for example in sound filters♣ (also called tone controls, or EQ). Non-linear distortion occurs in a non-linear system. For a system to operate linearly, it needs to have proportionality, lack of sources, and the possibility for superpositioning. If only one of these conditions is missing, the respective system is non-linear.

- "Proportionality" means that an n-fold increase of the input signal will result in an n-fold increase of the output signal (doubling the input results in doubling of the output).
- "lack of sources" means that without an input signal, the output signal will be zero.
- "possibility for superpositioning" means that a transformation of a sum of signals effected by the system will correspond to the sum of the transformed individual signals :$T(x+y) = T(x)+T(y)$.

The description will be simpler if linear and non-linear systems occur strictly separately, i.e. if every sub-system will perform only linear or only non-linear mapping. A non-linear system that does not cause any linear distortion includes no memory – this is because only devices including memory (inductances, capacitances) generate frequency-dependent resistances and thus create a frequency-dependent transmission. In a **memory-free system**, the output signal will therefore not depend in any way on the past input signal but exclusively on the input signal occurring at the very same moment. The transmission characteristics can be described via a characteristic $y(x)$. For non-linear behavior to be present, we need to have a curved characteristic (strictly speaking, an offset also introduces non-linearity).

The amplifying elements (tube, transistor) used in guitar amplifiers all feature a curved characteristic, and therefore every guitar amplifier operates as a non-linear device. According to the rules of classical amplifier technology, these non-linearities are supposed to be as small as possible, and therefore negative-feedback circuits reduce the gain and at the same time perform a linearization. Many guitar players were satisfied with the resulting so-called "clean" sound, but some forced non-linear distortion by overdriving their amplifiers, creating "crunch", "distortion", or "fuzz". In many amps this required using their full power and thus very high loudness – but sometimes even at maximum gain, the resulting non-linear distortion was not pronounced enough. This is the reason why additional devices for the generation of non-linear distortion were created under various monikers: fuzz-box, distortion-pedal, overdrive … Before long, the effect was not limited to additional devices: with an increasing number of tubes, guitar amplifiers themselves soon offered possibilities to control the desired degree of distortion.

The distortion devices described in the following are systems that add **non-linear distortion** to the guitar sound. Whether this happens in the amplifier itself or in a separate device is not distinguished to begin with. With the distortion, the guitar sound becomes fuller, sustaining, more shrill, more aggressive, buzzing, more alive – this is always depending on the chosen settings. There is not "the" distortion sound. The distortion also changes the **dynamics** of the signal in the sense that sustain is extended. Since practically all distortion devices have a degressive, limiting characteristic curve, any level-differences in the guitar signal are reduced and differences between loud and soft are evened out. The originally percussive guitar sound becomes steadier, and takes on some sound-characteristics of horns (saxophone, trumpet) or strings (cello).

♣ M. Zollner: Signalverarbeitung. Hochschule Regensburg, 2009.

There are two types of drive-situations for customary tube- and transistor-circuits: low drive-levels with weak distortion, and overdrive with strong distortion. In the left section of **Fig. 10.8.20**, we see the characteristic curve of an ECC83, and in the right-hand section the time-function of the sine-shaped input voltage and the distorted output voltage. For small drive-levels (up to about 1 V_S), there are only small differences between input- and output-voltage. For high drive-levels, we find a strong – and in this case asymmetrical – limiting of the signal ("clipping").

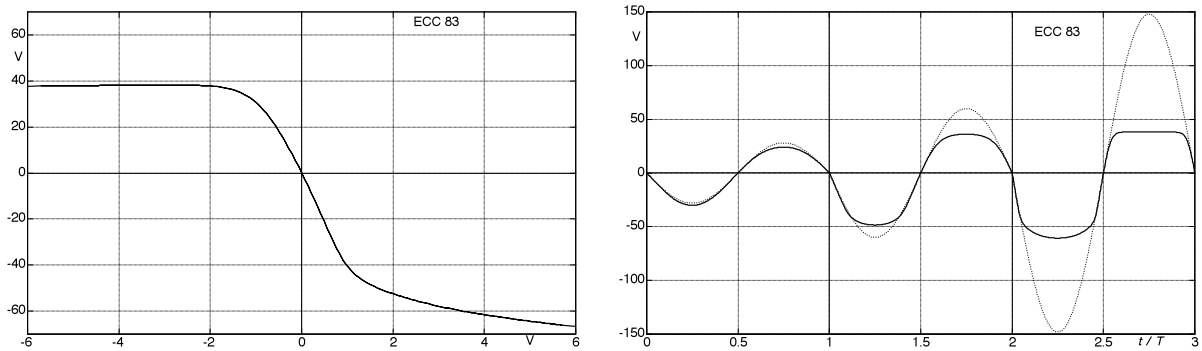


Fig. 10.8.20: Non-linear distortion for an ECC-83 (compare to Chapter 10.1.4).

Now, the human auditory system is not a sensor for directly analyzing the time-function. Rather, it detects in a first step the time-variant short-term spectrum (Chapters 8.2.4 & 8.6) to determine the sound color. If a sine-tone (e.g. 1 kHz) undergoes non-linear distortion, new spectral lines are created at the integer multiples of the fundamental frequency, i.e. at 2 kHz, 3 kHz, 4 kHz, etc. Distortion of a signal composed of several partials will generate sum- and difference-tones as multiples of the largest common divisor of all partials♣. If the sound of a single guitar string were of strictly harmonic content, distorting it would still result in a harmonic spectrum. The level and the phases of the partials would change, and so would the sound color, but the frequency of the partials would remain unchanged. However, the spectrum of every real string-vibration is spread **in-harmonically**, and it is here where we find the key to understanding the impact of a distortion device.

If, for example, a complex tone consisting of a 100-Hz- and a 202-Hz-partial undergoes 2nd-order distortion, additional partials at 0 Hz, 102 Hz, 200 Hz, 302 Hz, 404 Hz are created. The 0-Hz-component may be ignored because it is DC that the circuit will not transmit further. The partials at 302 and 404 Hz will brighten up the sound if they are strong enough, but the main effect will be close to the primaries: the partials at 100 and 102 Hz will beat against each other, and so will the 200- and 202-Hz-partials. The non-linear distortion will, on one hand, enlarge the spectrum towards high frequencies (i.e. emphasize the treble more), and on the other hand the amplitudes of the primaries will start to fluctuate due to the beat effects: the sound becomes more lively. If not only 2nd-order distortion occurs but higher-order distortion as well, a large number of additional partials is created and correspondingly many and possibly strong fluctuations. These fluctuations bring a kind of noise-character to the sound; we get an effect as if additional **noise** would be superimposed. Strictly speaking, noise in its usual definition belongs to the group of stochastic (random) signals, while the fluctuation of partials generated by non-linearity is not stochastic but determined. Since this special noise has a periodicity, it is called **pseudo-noise**.

♣ M. Zollner: Frequenzanalyse. Hochschule Regensburg, 2009.

Fig. 10.8.21 depicts the zero-symmetric characteristic curves that will be used for distortion in the following. They are odd-order functions, i.e. functions that can be expanded into a series containing only members of odd-order power (x, x^3, x^5, ...).

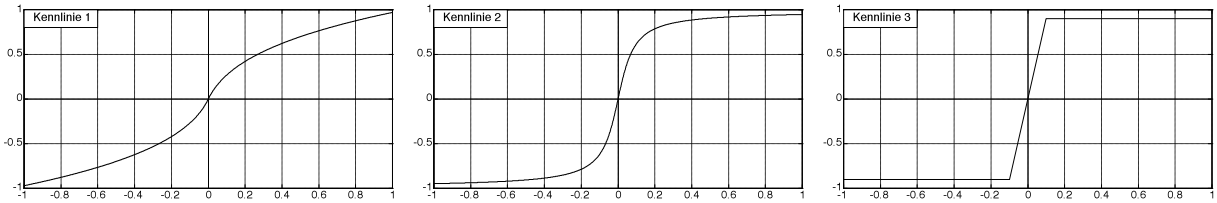


Fig. 10.8.21: Characteristic curves of the systems used for distortion.

Simple periodic functions are distorted in **Fig. 10.8.22** using characteristic curve 2 (an arctan-function). Distortion of the sine-signal (uppermost row in the figure) results in a spectrum with only odd-numbered harmonics. However, merely adding a second partial to the primary signal (2nd row in the figure) may generate even-numbered harmonics – although this is not generally the case, as demonstrated by the third row in the figure. Only a signal of half-wave anti-symmetry will, in its spectrum, contain exclusively odd-numbered harmonics. Such a half-wave anti-symmetric signal, if distorted via an odd-order characteristic curve, remain half-wave anti-symmetric, and will not gain any even-numbered harmonics. In the lowermost line of the figure, a signal of three partials is distorted. Due to the 2nd harmonic, this signal cannot be half-wave anti-symmetric, and therefore the spectrum of the distorted signal contains even-order harmonics as well.

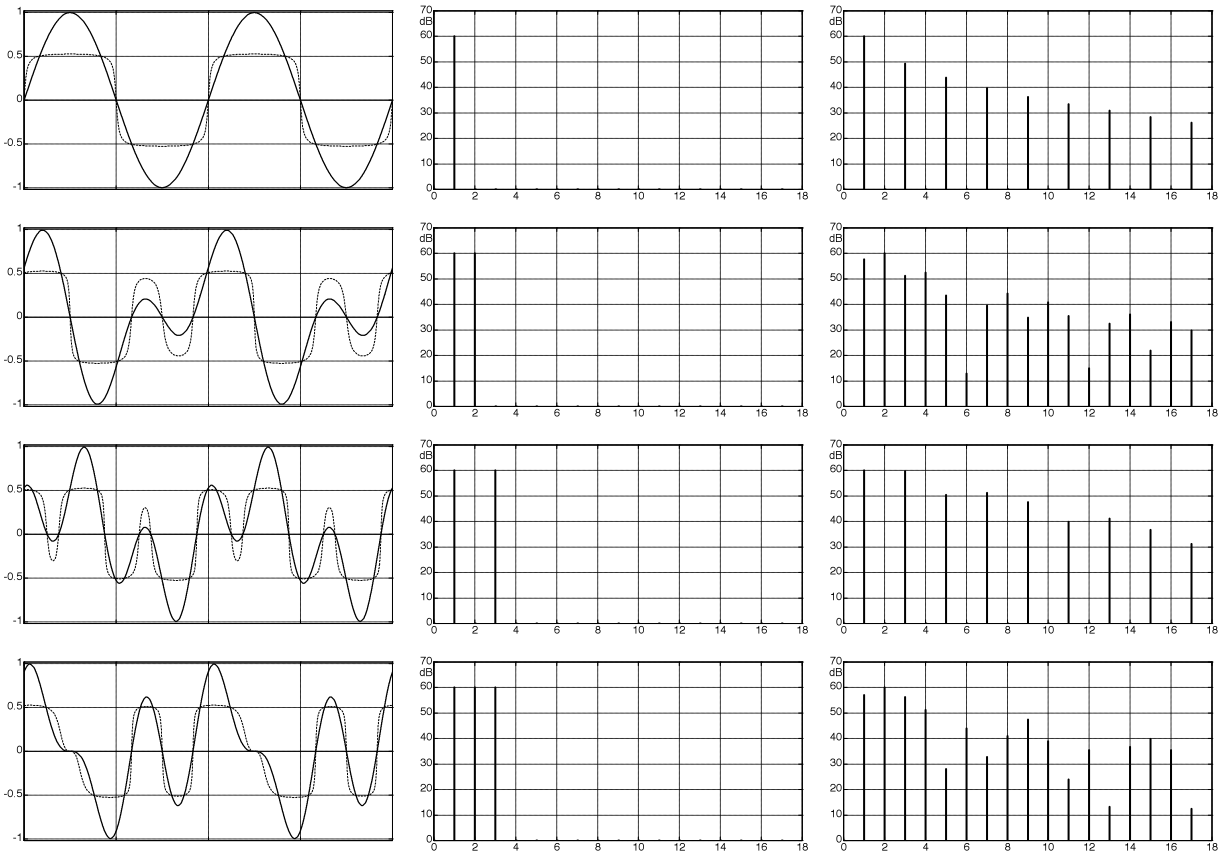


Fig. 10.8.22: Time-function of undistorted and distorted signal (left), spectrum of undistorted signal, spectrum of distorted signal (right). Time-functions and spectra individually scaled. Characteristic curve 2.

For the analyses discussed in the following, synthetic guitar sounds were used; their rules of construction were nonetheless derived from a real guitar. Synthetic sounds were used because all parameters of these sounds are known – which is not the case for real sounds. For the **harmonic** signal, 45 partials were added. The frequencies of these partials all had an integer relationship to the fundamental of 82 Hz, and their levels and decay time constants were taken from a real guitar tone (open E_2-string). The **inharmonic** signal was synthesized with the same levels and time constants, but the frequencies of the partials were slightly spread out according to the formula outlined in Chapter 1 (dispersion due to the bending stiffness of the strings). The spectrograms of the undistorted and distorted synthetic guitar sound are shown in **Fig. 10.8.23**. The spectrum of the harmonic sound ends at 3,7 kHz in view of the usage of 45 partials; the spectrum of the inharmonic sound goes up to 4.1 kHz due to the frequency spreading (again 45 partial were used).

Distorting the **harmonic** sound results in additional partials that however are positioned exactly within the harmonic grid. On one hand, the new partials fill up the frequency range above 3.7 kHz, and on the other hand they change the level of the primary partials. The degressive curvature of the distortion characteristic has the effect that the partials decay more slowly, for some there is even an initial growth. The changes of the partial-levels are slow, with change speeds similar to those of the primary levels. For the **inharmonic** sound, the distortion generates many new partials positioned closely to the primaries such that the DFT-analysis (and the auditory system) cannot recognize them individually anymore. The spectral pooling of these undistinguishable lines results in fast signal modulations bearing some resemblance to a stochastic noise process but being (strictly speaking) determined (pseudo-noise).

The distortion has three effects on the sound: the treble-content grows (a more brilliant sound), the dynamics are compressed (longer sustain), and the partials are pseudo-stochastically modulated (creating a “buzzing” and “raspy” character). The pseudo-stochastic modulation happens only for inharmonic sounds and is dependent on the **string-parameters**. **The thicker the string, the more noise is created**. Maybe we should explicitly mention that this holds for single tones, because for chords, the spectrum is not harmonic in a simple fashion anymore, anyway.

The **level evolutions** shown in the picture below indicate that the pseudo-stochastic modulations increase if the characteristic is more strongly curved. Analyses for characteristic 3 are not included; a similar picture as for characteristic 2 would emerge, as long there is strong overdrive. Larger differences become apparent for less overdrive: the change from “undistorted” to “distorted” happens abruptly for characteristic 3, and more gradually for the other characteristics. Moreover, the spectrum becomes more treble-heavy if abrupt signal-limitation (clipping) occurs.

It should be expressly mentioned here again that despite a zero-symmetric characteristic (“odd-order function”), distortion products of even-order do occur. The assumption, that an odd-order-characteristic would generate only odd-order distortion products, only holds for half-wave anti-symmetric signals (e.g. for a sine tone). For real guitar sounds, even-order distortion products can very well result from odd-order characteristics. In the same manner, the assumption that tubes would generate predominantly even-order distortion products is wrong as a general statement. Tubes do not generate “better” distortion than transistors – otherwise nobody would ever have used a Range-Master ahead of the amp. In the Range-Master, pure transistor distortion is generated (Chapter 10.8.5.3).

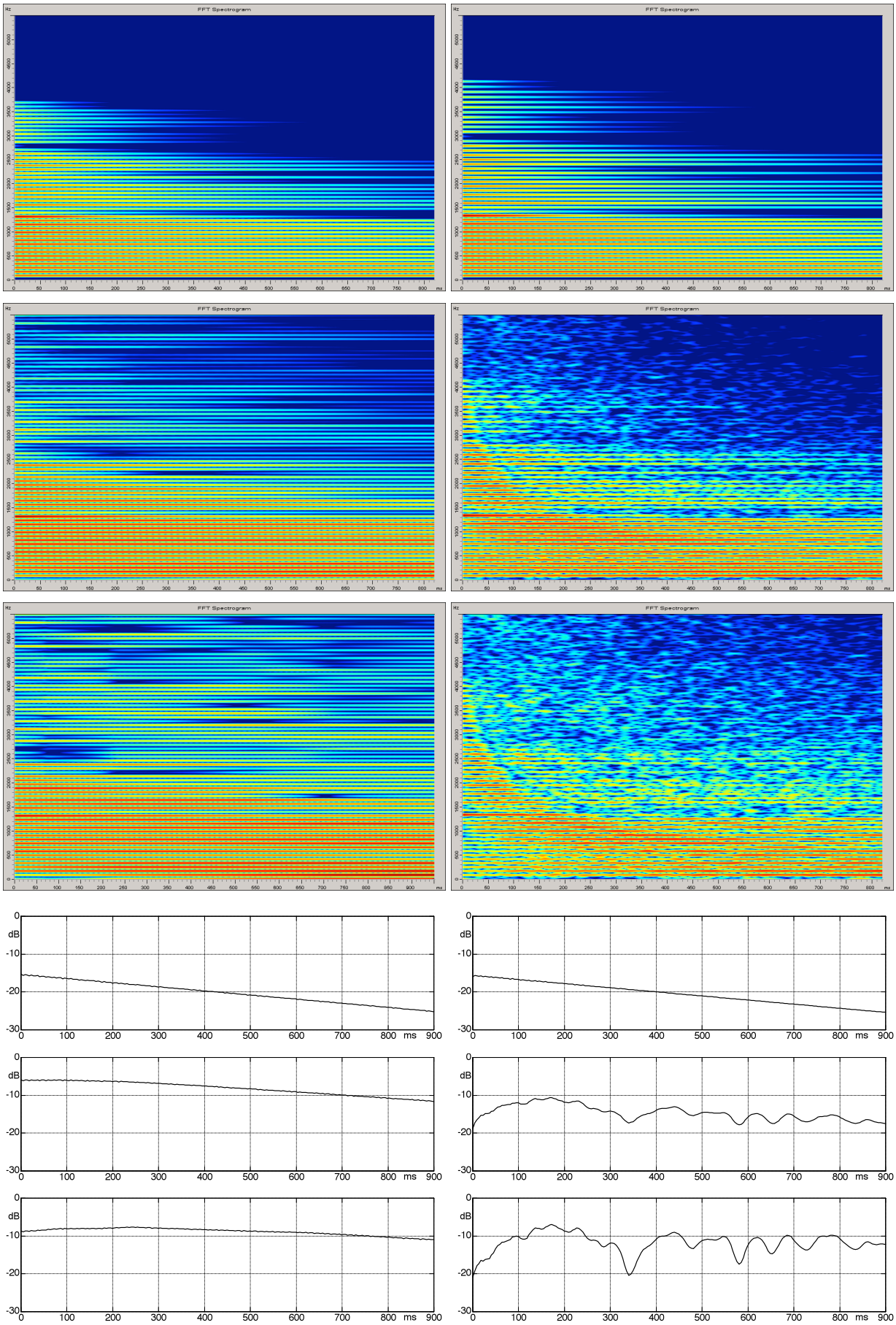


Fig. 10.8.23: Spectrograms, harmonic (left) und inharmonic signal, 0 - 5.5 kHz, ΔL = 40dB.
1st row: no distortion, 2nd row: characteristic 1, 3rd row: characteristic 2. Bottom: level-evolution of 15th partial.

Digital emulations of analog distortion devices merit some special consideration. It is generally known that an anti-aliasing filter needs to be included ahead of an A/D-converter in any time-discrete system. This is in order to avoid spectral back-convolutions (sampling theorem). If the signal is already appropriately bandwidth-limited, this filter may of course be dispensed of. For example, there is no need to include a filter if a 5-kHz-tone is sampled at 44.1 kHz. If the A/D-converted 5-kHz-tone is, however, distorted in the digital realm with a digital distortion characteristic, new frequency lines are generated – above half of the sampling frequency, as well. Since time-discretization has the effect of spectral periodization, these new lines appear around all multiples of the sampling frequency. If the distortion characteristic is point-symmetrical, new partials are generated at 15 kHz, 25 kHz, 35 kHz and further odd-numbered multiples. The distortion products are mirrored with respect to the sampling frequency (e.g. 35 kHz re. 44.1 kHz), a new distortion line appears at 9.1 kHz that would not be generated by an analog distortion device. At 0.9 kHz, and at many other frequencies, further partials appear, sounding rather unpleasant (as a rule, if their level is high enough). It is therefore insufficient to digitally emulate an analog distortion characteristic in order to create a digital equivalent. The higher the frequency of the signal to be distorted, and the more angularly shaped the distortion, the more disappointing the emulation will be.

To avoid such back-convolutions, the sampling frequency needs to be increased. Whether a ten-fold increase is adequate or whether even much higher sampling rates are necessary, depends on the signal, the distortion characteristic, and the quality requirements. Here is a simple estimate: if a sine signal undergoes hard clipping with a symmetric rectangular characteristic, new partials are generated following an si-envelope. The level of the 11th partial is 21 dB below the level of the primary, the 99th partial is 40 dB below the level of the primary. If the sampling frequency is 100 times of the frequency of the tone to be distorted, the back-convolution creates an interfering tone which is 40 dB down relative to the primary-level (**Fig. 10.8.24**). Of course, not only this one interfering tone is back-convoluted, and back convolution does not happen only at the sampling frequency. The figure shows merely one back-convolution such the lines can still be associated properly.

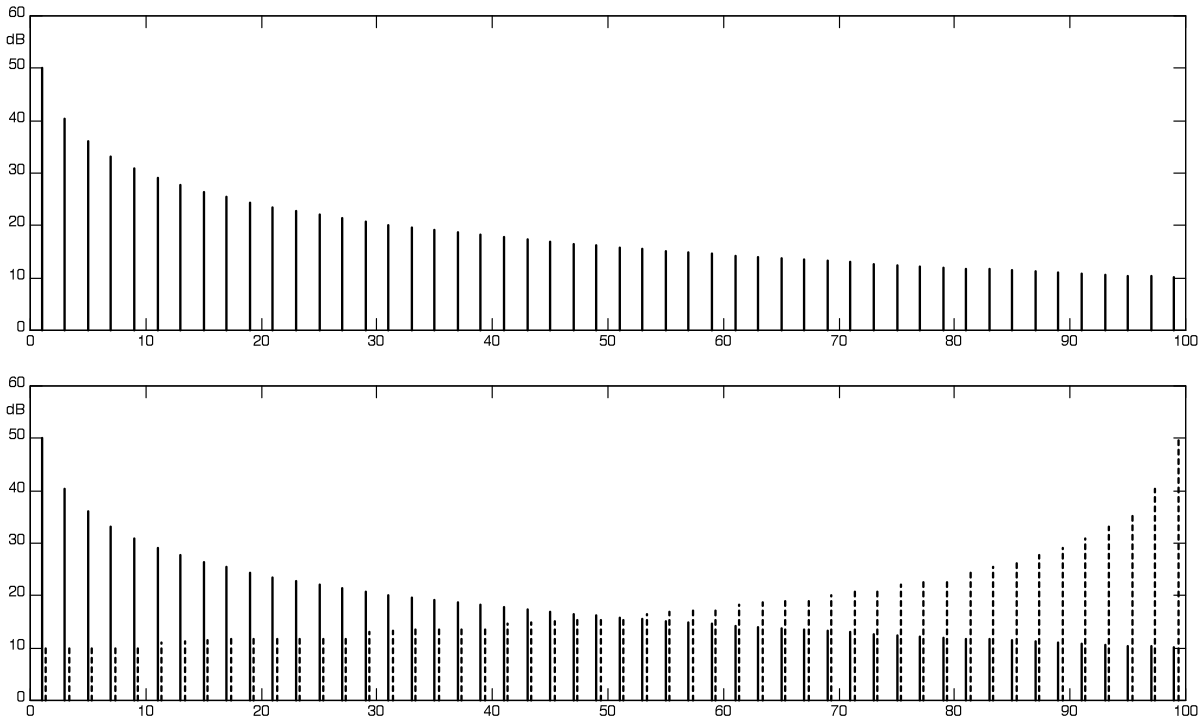


Fig. 10.8.24: Spectrum of a strongly distorted sine tone; time-continuous (top); time-discrete with one back-convolution (bottom). The frequency of the back-convoluted lines is strongly dependent on the relation f / f_a .

10.8.5.1 Diodes

In order to achieve non-linear distortion of a signal, at least one non-linear component is required. This may be a tube, a transistor, or – in the simplest case – a diode. In the following, the term diode is meant to refer to a **semiconductor diode** and not to a tube diode. The latter could also be used for distortion but this does not happen in practice. Simplified, diodes conduct current only in the forward-direction; with the reverse-polarity, the diode blocks. More accurate models consider that a few 100 mV are created across the diode in the forward-direction, and moreover include the reverse-current flowing for reverse polarity. Dynamic models add capacitances (possibly of non-linear nature) – a model for a diode can in fact get quite complicated. As a first approximation, the so-called Shockley-characteristic suffices:

$$I = I_S \cdot \left(\exp(U/U_T) - 1\right); \qquad U_T = 25.8\,\text{mV}\ (300\,\text{K}) \qquad \text{Shockley-equation}$$

The diode current I grows exponentially with the voltage across the diode in the forward-direction U which is referenced to the temperature-voltage U_T; I_S represents a theoretical reverse-current. Real diodes may strongly deviate from this idealization, and corrections and supplements are necessary. In particular, it is necessary to modify the temperature-voltage: for a real diode, measurements yield values for U_T of up to more than 60 mV; moreover a track-resistance in forward-direction needs to be considered. In the left section of **Fig. 10.8.25**, the forward characteristic of an 1N4148 diode is shown with linear scaling, the middle section shows the same characteristic but with log-scaling along the horizontal axis. The exponential function leads to a strong curvature for the linear scaling – this led to the term "threshold voltage"; for silicon diodes this is often specified at 0.7 V. A scaling for large currents indeed shows a sharp bend of the rounded characteristic at 0.7 V; a scaling for smaller current shifts the "kink" to 0.4 V or even smaller values. Do note: an exponential function does not have a kink – the value of the threshold is arbitrary!

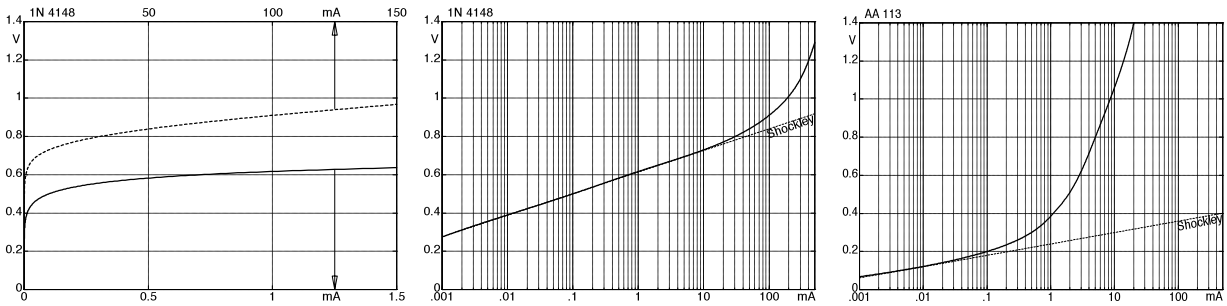


Fig. 10.8.25: Pass-characteristic of a 1N4148 (data from Fairchild). The upper x-axis-scaling in the left picture holds for the dashed line, the lower scaling for the solid line. Right: AA113 (Siemens).

The right-hand picture shows the forward characteristic of a Germanium point-contact diode (AA113). In contrast to the silicon diode 1N4148, smaller voltage occur across the diode for small currents in the forward-direction. For currents above 3 mA, however, the voltage across the Ge-diode is higher than that across the Si-diode because spreading resistances make themselves felt more.

For all these characteristics, we need to bear in mind that there will be scattering due to temperature- and manufacturing-fluctuations. Increasing the temperature by 20°C may indeed double the forward-current, and exchanging a diode for another of the same type (!) may change the current flowing at a given voltage within a range of -70/+200%. It is therefore not purposeful to count on highly specific data from the data sheets.

Fig. 10.8.26 shows further forward-characteristics (from data sheets). All diodes marked AA are Ge-diodes, all others are SI-diodes.

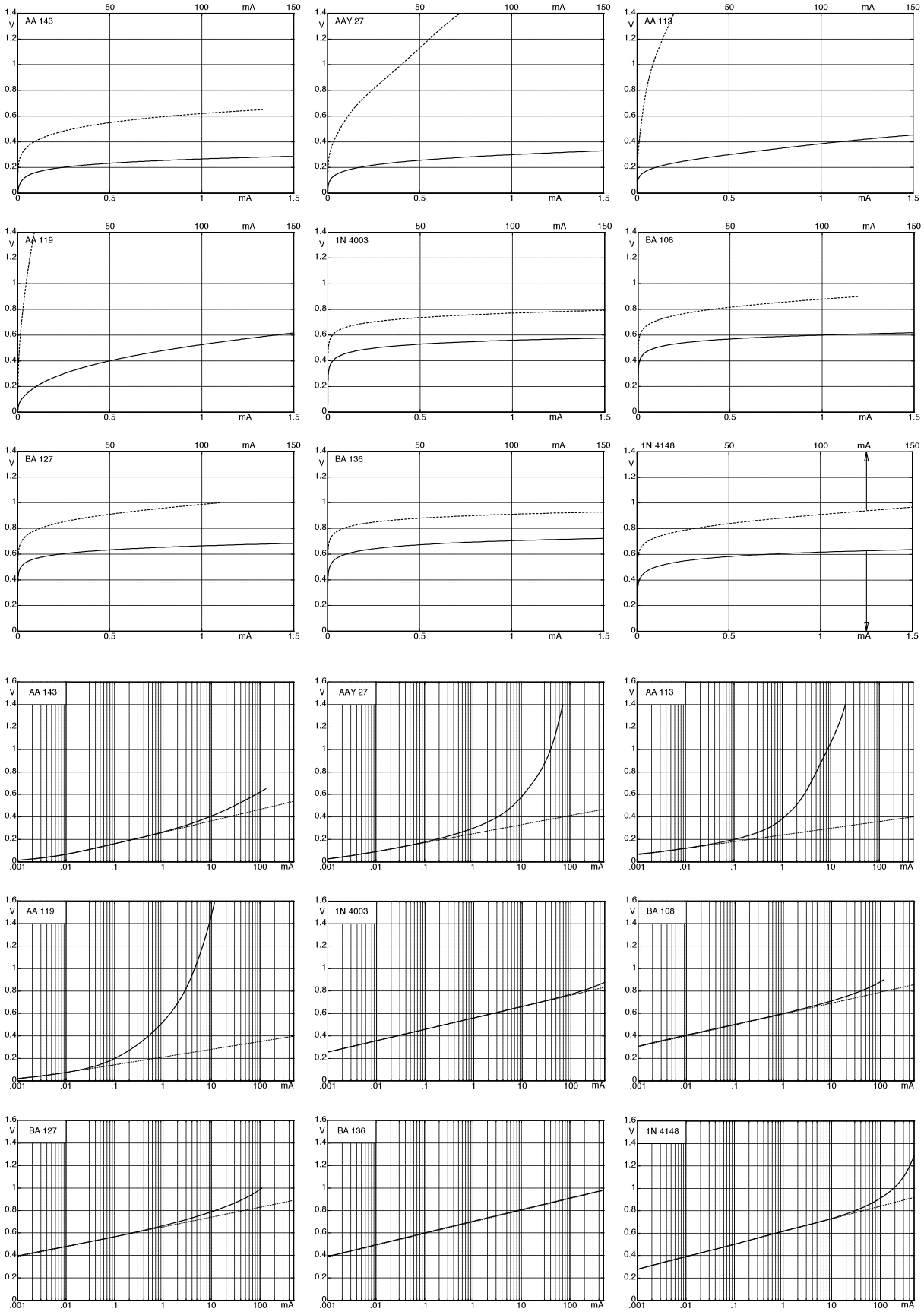


Fig. 10.8.26: Forward characteristics of various semiconductor diodes (data sheet specifications).

The diode currents flowing in typical circuits for distortion devices are small (max. about 1 mA). The characteristic curves (reaching over 100 mA in the data sheets) are therefore only relevant in the lower range – which is relatively well described via the Shockley-equation. To achieve a limitation of both sides of the signal, two diodes are interconnected in an anti-parallel fashion; this results in a zero-symmetrical (odd-order) characteristic. In theory, that is: manufacturing tolerances are felt considerably in real diodes. Driving the diode-pair by a stiff current source results in a signal rounded off at both sides, as depicted in **Fig. 10.8.27**). This ideal-current-source-drive is, however, not possible (and not necessary) in reality, since the impedance of the driving current source cannot become infinite. It is helpful to use, for the model, an ideal current source and extend the diode pair by a parallel resistor (left-hand picture) – this changes the behavior in particular at small drive-levels. In all 3 pictures we see the characteristics of two diode pairs: for the small-signal diode 1N4148, and for the power diode 1N4003.

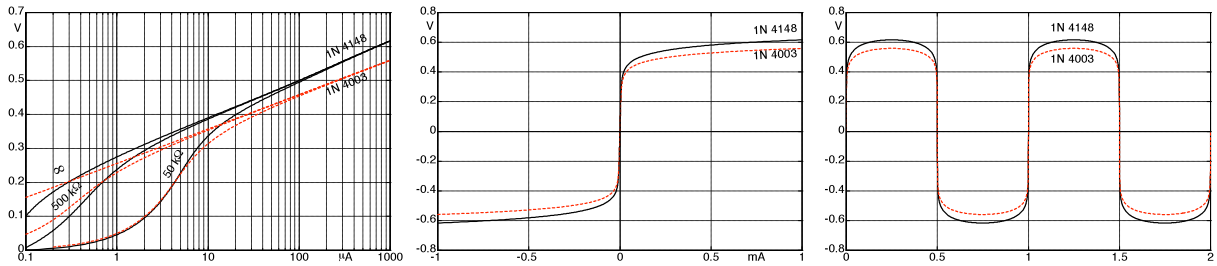


Fig. 10.8.27: *Left:* forward characteristics (half-log representation), diode with resistor in parallel. *Middle:* two anti-parallel diodes. *Right*: voltage limiting for sinusoidal current input.

Describing the diode characteristics with only the Shockley-model requires merely **two parameters**: the reverse current and the factor of the temperature-voltage (1…2). Both parameters can be seen as scaling factors for the current and the voltage, and consequently the following holds: within the framework of the Shockley-model <u>all</u> diodes show the same behavior as long as variable gain is available at both the distortion device input and the distortion device output. Working with this model does not require choosing a special diode because every diode allows for the same distortion characteristics. However, this does not tell us anything about the **dynamic** behavior, which is not described via a static characteristic curve. A diode will go into the blocking state only once "all" charges have left the barrier layer output, and this takes a moment: a relatively long moment for power diodes and a relatively very short moment for RF-diodes. Partnered up with the distortion-introducing diodes operating within the feedback branch of an operational amplifier, we often find an additional capacitor in parallel to the diodes; this leads to the conclusion that it is in fact not even desirable that the diodes act very quickly. In the **Tube-Screamer**, for example, there is a 50-pF-cap in parallel that will have an effect only in the highest frequency range – and only with the gain turned up. In the **Boss DS-1**, however, we find some quite respectable **10 nF** in parallel with the diodes! This rather huge capacitance pushes the switching behavior of the diode somewhat into the background. Moreover, we must not forget that even a capacitor of this size impresses a diode only as long as very small currents are flowing. At e.g. 2 mA (a current value that certainly may occur), the differential resistance of the diode is a mere 20 Ω, and compared to that even 10 nF are of relatively high impedance.

Different diodes are connected in an anti-parallel manner if a non-zero-symmetrical characteristic is desired – e.g. a Ge- and a Si-diode, or special parallel/serial-networks. The individual characteristic becomes more important in this scenario, because it is not possible to do an individual current/voltage-scaling anymore for each diode.

10.8.5.2 Transistors

With transistors, we need to distinguish between two crystal types (Ge, Si) and two types of doping (PNP, NPN). Strongly simplified, the main differences♣ are: NPN-transistors require (in common-emitter configuration) a positive operating voltage, and PNP-transistors require a negative one. The typical base-emitter-voltage is 0.1 V for a Ge-transistor and 0.6 V for a Si-transistor. There is a vast multitude of the most different transistors – and among them a surprising number of compatible equivalents. PNP vs. NPN is, however, incompatible, as is Ge vs. Si, even though there may be instances where the latter swap will work. An OC44 may be exchanged for an AC151 without any problem, but an AC187 is incompatible with an AC188. For not too old specimen of Europe-built transistors, the first letter in the designation specifies the crystal-type: A for Germanium, B for Silicon. The second letter stands for the recommended usage: C for audio frequency preamplifiers, D for audio frequency power amps, F for RF-amplifiers, S for switching stages. The American (2N) and Japanese (2S) designations do not allow for such a distinction.

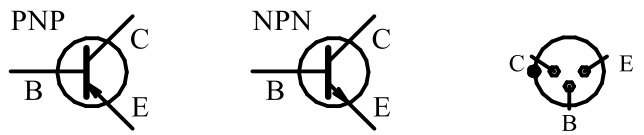


Fig. 10.8.28: Transistor-schematics, connections.

Fig. 10.8.28 shows the circuit diagram for transistors, and the connector-pin assignment (seen from below). Not all transistors have this assignment – in case of doubt the data sheets of the manufacturers help. For an **NPN-transistor**, the current flows into the base, out of the emitter and into the collector (the technical direction of the current-flow), and for the **PNP-transistor** out of the base, into the emitter and out of the collector. The usual collector-current values in distortion devices are smaller than 1 mA; the base current is about 1% of the collector current. The quotient of collector current and base current, i.e. the **current gain** B, is strongly dependent on the manufacturing process and the temperature. For commonly used transistors, B is about 40 … 300. It is therefore possible that the behavior of a circuit changes if one transistor is swapped for another (of the same type!).

In the idle state (i.e. without input signal) the collector current is about 0.1…1 mA. This of course depends on the specific circuit – as example we assume it to be 0,2 mA. For $B = 100$ the base current will amount to 2 µA. The input, i.e. the "gate" between base and emitter, is best described as a diode operated in the forward-direction – a Si-diode for the Si-Transistor, and a Ge-diode for the Ge-transistor. A forward-current of 2 µA yields a forward-voltage of about 0.1 V for the Ge-diode and of about 0.5 V for the Si-diode. Again, this is a first point of reference – depending on the manufacturing process these values may vary. If the base-voltage for an NPN-Transistor is more than 1 V larger than the emitter-voltage, that transistor is shot. If U_{BE} is negative for an NPN-transistor, the transistor will be in blocking mode, and the collector-current will be approximately zero. The same correspondingly holds for a PNP-transistor and negative base-emitter-voltages. The collector-current will, however, not be exactly zero since a **reverse current** will still flow – in Ge-transistors this can reach sizeable values. For example, the Siemens data-book specifies a reverse current of max. 200 µA for the AC188 (for the emitter-diode in blocked state) – corresponding to the current in the operating point for the above example! In addition, the reverse current has the unpleasant characteristic of exponentially growing with increasing temperature. All this has created in particular for Ge-transistors the image that they are solitary, hard-to-handle lone wolves.

♣ For practice-oriented details see e.g. Tietze/Schenk: Electronic Circuits – Handbook for Design and Application; Springer.

In **Fig. 10.8.29** a PNP-transistor is operated in common-emitter configuration, i.e. with the emitter connected to ground. If the transistor is in blocking mode, there will be (almost) no collector current and the collector voltage will be -9 V (left picture). If the transistor is fully on, there will be only a small voltage left at the collector of e.g. -0.2 V. As a first approach, an operating point in the middle of the characteristic curve would be selected; the collector voltage would be set to -4.6 V. From this, we obtain a voltage across the collector resistor of 4.4 V resulting in a collector current of -0.44 mA for a resistance of 10 kΩ♣. The base voltage would be -0.1 mV in this example.

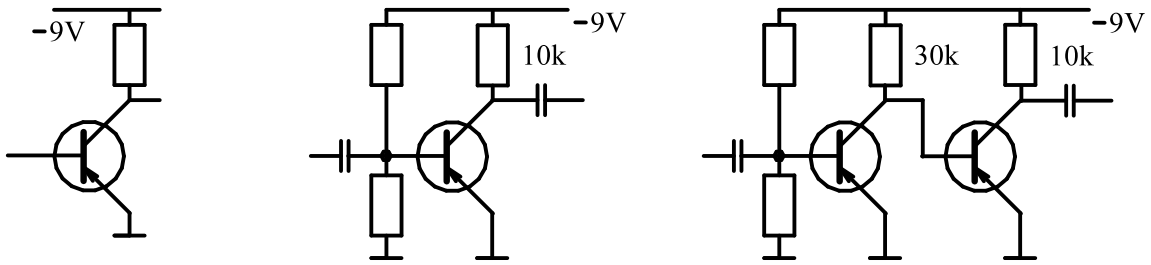


Fig. 10.8.29: Transistor in common-emitter circuit.

Both base and collector are not at 0 V without signal input, and a coupling capacitor each is necessary for connection to the dc-free outside world. The base voltage required for the operating point is set via the voltage divider at the base (middle picture). This circuit would not support a stable operation, however, even if the operating point would be set individually for each transistor specimen. With just a few degrees of temperature drift, the operating point would shift, and the sound would change. The means of choice countering thermal drift is **negative** (i.e. inverse-phase) **feedback**. This is implemented either via an emitter-resistor (increasing the input impedance), or via a resistor from the output back to the base (lowering the input impedance), or via other measures too extensive to be covered in the present context [see e.g. Fliege]. The following pages will show examples of transistor-circuits employing negative feedback – see e.g. chapter 10.8.5.3. Only with purposeful negative feedback, multistage amplifiers such as the one in the above right-hand picture can be put together. In the version shown, the first transistor would have to operate with too small a collector voltage: since the base voltage of the second transistor can not grow above about 0.2 V, the collector voltage of the first transistor is subject to the same limitation. This is why a resistor (of e.g. 1 kΩ) is introduced into the emitter branch of the second transistor; this resistor increases the input impedance and the input voltage.

Negative feedback decreases the gain but also stabilizes it, i.e. it becomes less sensitive to fluctuations in temperature or due to manufacture. Circuits that need not operate down to a frequency of 0 Hz allow for a separation of AC- and DC-negative-feedback. A strong negative feedback for DC will stabilize the operating point, while at the same time a weaker negative feedback for AC will ensure that the gain does not drop too far. One thing that needs to be considered for all amplifiers is the phase-shift that occurs at high frequencies: it can turn negative feedback into a positive one: the circuit may start to oscillate and inadvertently become an RF-generator.

♣ The in fact quite important area of reference arrows and algebraic signs will not be elaborated upon in this context – reference is made to literature, e.g. [20].

10.8.5.3 Range-Master (Dallas Arbiter)

One of the famous distortion devices from the early days of hard rock is the "Range-Master" built by the Dallas-Arbiter-company in Britain. It was also called a **treble-booster**, although it did not just increase the gain at high frequencies (as it happens when a "Treble"-knob is turned up) but performs this job in a non-linear fashion, with a rich seasoning of distortion. The circuit of the Range-Master (**Fig. 10.8.30**) is a simple as they come: the input signal is fed via a relatively small 6.8-nF-capacitor to a transistor providing an amplification of a factor of approximately 60. Since normal pickups generate several 100s of mV (even up to 4 V are not impossible), this transistor is almost always overdriven. However, two peculiarities need to be considered already at the input (U_1): The input impedance is – at about 10 kΩ – rather low, and for this reason the coupling capacitor has the effect of a strong attenuation of the low frequencies. To calculate the cutoff frequency, it is not only the input impedance of the range-master that needs to be taken into account, but also the pickup impedance that is part of the mesh. 6.8 nF and 10 kΩ would result in a cutoff frequency of 2.3 kHz; depending on the pickup this value drops to 1 – 2 kHz.

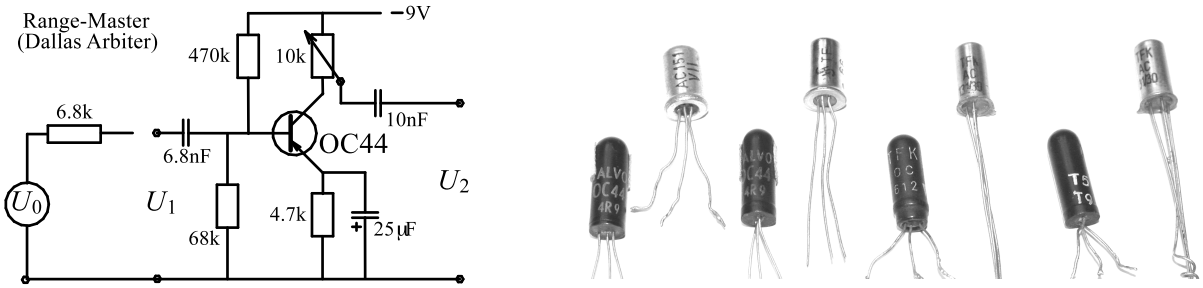


Abb. 10.8.30: Range-Master: circuit (left), old Germanium transistors (right).

The **OC44** used in the Range-Master is a Germanium-RF-transistor from the dawn of solid-state technology, and was probably available at very low cost – deemed outdated already back in the day given the high-speed progress in technology. Even in fully conducting mode, the collector current remains below 1 mA (for normal loads); the quiescent current flowing through the output potentiometer is merely 0.2 mA. The corresponding base current is 1 – 2 µA, the base-emitter-voltage is smaller than 100 mV. The quotient of collector- and base-currents (the current-gain $B = I_C / I_B$) has a large scatter-range due to manufacturing tolerances; values of 50 … 200 are possible. Thus, the operating point is also subject to scatter: typically, we find -6.8 … -7 V at the collector. For a new battery, that is – the power source will also influence the transmission behavior.

Connecting the Range-Master to an amplifier input of 1 MΩ impedance, the output capacitor (10 nF) of the Range-Master creates – in conjunction with this load – a high-pass with a cutoff frequency of 16 Hz. Choosing the low-sensitivity input of the amp (typically 136 kΩ input impedance) pushes the cutoff frequency up to 114 Hz. In total, **two high-passes** have an effect: the first at the Range-Master-input, the second at its output. The emitter resistor is so effectively bridged for AC that the resulting high-pass may be ignored: it creates no attenuation of the bass frequencies. It may not be ignored, however, regarding its effect on shifting the operating point. This happens when an input signal is present because now charge reversals of the emitter capacitor take place. Due to the non-sine-shaped emitter-current, the emitter-voltage shifts by about 0.2 V towards the negative at high drive-levels. Similarly, the polarization voltage of the input capacitor changes: the asymmetric base current flowing at overdrive-levels decreases the average voltage at the input cap. These shifts in the potentials are the "secret" of the Range-Master – not the purportedly unique behavior of the OC44.

An original OC-44 may be microphonic and/or be very noisy, its reverse current may be beyond good and evil – still it is being traded at prices 50 times of what modern transistors cost. What about the unique sound? That can be achieved with other transistors, as well. Of course, the latter need to be PNP-Germanium transistors, and an individual check is warranted for modern merchandise, too. An original OC-44, however, does not need to be put into the Range-Master – the operational behavior of different transistors can, in fact, be surprisingly similar. We may find large differences in the maximum ratings; for example with the collector voltage: 20 V, or 100 V. The same for the maximum collector current: e.g. 10 mA, or 2000 mA. And of course for the β-cutoff frequency: 150 kHz, or 10 kHz. However, all these values are of secondary importance for a distortion device operated at 9 V. Important is the current gain (static and dynamic) and the reverse current – and both these are not in any way special for the OC-44. Why else could B. C. Meiser♣ recommend as replacement the AC122, or the AC128, or – particularly suitable – the AC151. It should be noted that this recommendation is not the writing of a blind amateur (as it often is the case in magazines), but the well-founded opinion of a seasoned, experienced circuit designer. N.B.: the **AC128** is recommended in the data sheets "for slow switches" or for "small audio power stages", the **AC122** for audio preamplifiers, and the **AC151** for audio-preamps and driver stages. The **OC-44** was designed as RF-transistor for AM-radio usage … and still it can be swapped for these other transistors. Of course, there are piles of other possible transistors – the special sound is not due to a special transistor, but due to the asymmetric and drive-dependent transmission characteristic (which in itself is not even that unusual).

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Fig. 10.8.31: Range-Master: transmission characteristics; $U_0 \Rightarrow U_1$ (left), $U_1 \Rightarrow U_2$ (right).
This figure is reserved for the printed version of this book.

Fig. 10.8.31 shows, in its right-hand part, the transmission behavior from input (U_1) to output (U_2); the operating point is indicated via the coordinate axes. For small drive-levels, the correspondence between input- and output-voltage is approximately linear (short, straight section of the characteristic); with increasing amplitude the characteristics grow longer and curved, and shift to the left towards more negative input voltages. As already mentioned, the reason for the shift is the charge reversal in the capacitors. Without the shift, limiting on both sides would occur already at U_1 = -0.2 V; with the shift it happens only at U_1 = -0.4 V. Put another way: the shift of the operating point renders the characteristic less symmetric and emphasizes even-numbered distortion (for a sine-input). Another effect is the dependency of the operating-point-drift on the input signal amplitude: it changes transmission parameters more strongly than a fixed characteristic: the guitar-sound increases in liveliness. To emphasize it again: all this is not a special OC-44-characteristic – every suitable transistor will take care of these effects. But know this: many are suitable, but few are chosen …

♣ Gitarre&Bass, 01/2002.

In the left section of **Fig. 10.8.31**, the correspondence between U_0 and U_1 is show. A voltage source was connected via 6.8 kΩ to the input of the Range-Master; the non-linear **input impedance** of the latter is the reason for the curve in the characteristic. For a (passive) magnetic pickup, the Range-Master represents a very special load: non-linear and of low impedance such that the pickup resonance cannot manifest itself. **Fig. 10.8.32** indicates the effect assuming linear filtering: the low-impedance load attenuates the treble upwards of about 1 kHz, the series capacitor attenuates the bass, and, from the pickup-source-voltage (in this case a Strat) to the Range-Master input, we obtain a **bandpass characteristic** with a center frequency of 1,25 kHz.

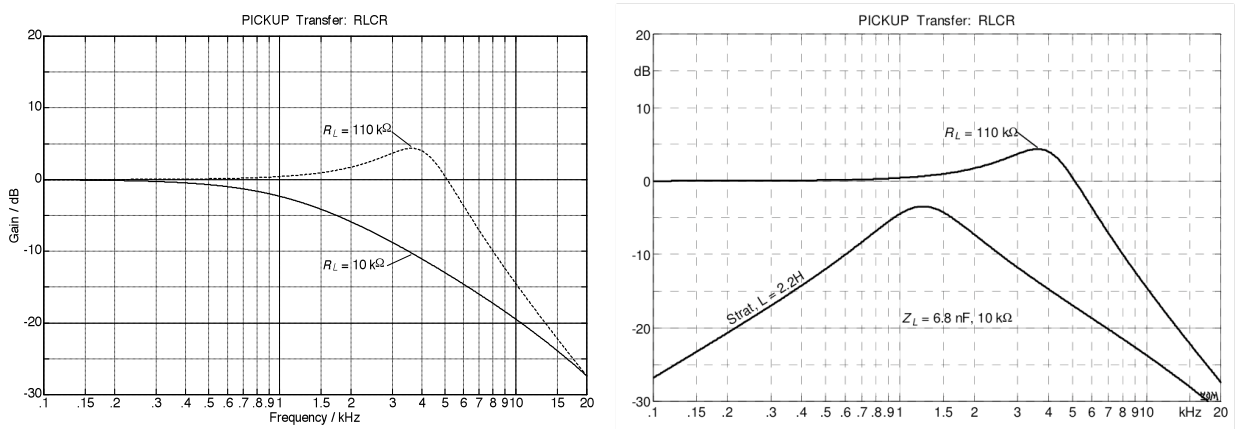


Fig. 10.8.32: Low-pass model of the loaded pickup (see Chapter 5.5.4). In the left picture, the Strat-pickup is subjected to a real load of 10 kΩ, in the right-hand figure the load is the series circuit of 6.8 nF and 10 kΩ.

Fig. 10.8.33 depicts the non-linear distortion for a sine-shaped input signal. The lower half-wave of the collector voltage is cut off first; the characteristic is not point-symmetrical and the duty cycle therefore is not 50%. These measurements were taken at 500 Hz, and the input capacitor was enlarged to 680 nF in order to be able to clearly separate linear and non-linear distortion. For the regular operation (6.8 nF), linear and non-linear distortion interact.

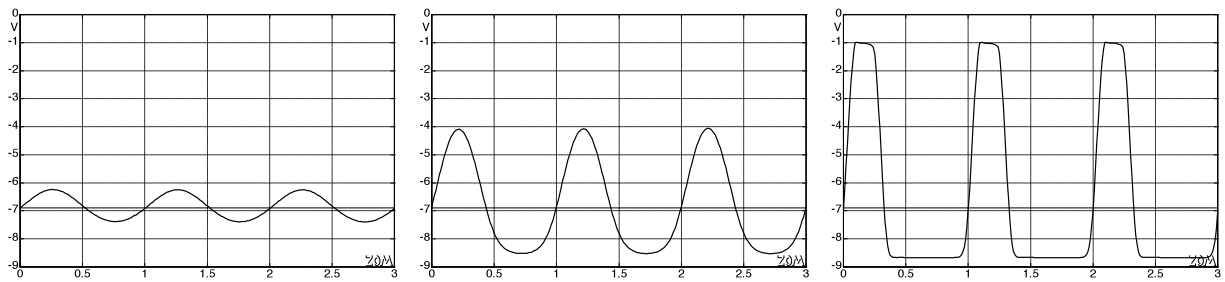


Fig. 10.8.33: Range-Master: collector voltages for sine-shaped input signal; f = 500 Hz.

In the following table transistors are listed that may serve as replacements of the OC44; the limit values are taken from data sheets (Va, Te, Si, and others).

	OC44	AC122	AC125	AC126	AC128	AC151	2N508	2N527	OC71	OC75	OC77
kHz	150	15	17	17	15	15	45	35	10	8	3.5
mA	5	100	200	200	2000	200	100	500	10	10	125
V	15	18	32	32	32	32	16	45	30	30	60

Table: transistors comparable to the OC44 [B.C. Meiser, Gitarre&Bass 1/02]. It follows from the data variance that practically every Ge-small-signal transistor is suitable; optimum-β = 80-110.

10.8.5.4 Tube-Screamer (Ibanez)

Whether you want to call the Tube-Screamer a distortion device, an overdrive or a treble-booster is a matter of taste – there is no fixed rule. The unit goes between guitar and amplifier, and it has 3 sections: an impedance converter consisting of a transistor, a distortion section with low-cut, and a tone filter. **Fig. 10.8.34** shows – grouped around an operational amplifier – the distortion-section components. A high-pass at the input takes care of a defined operating point – its cutoff-frequency is so low that it has no impact on the frequency response. The negative feedback circuit of the OP-amp includes the distortion-significant diodes. They have merely a negligible effect at very low signal levels; here, the circuit operates as an amplifier with a high-pass cutoff-frequency of 720 Hz. However, as soon as the voltage across the diodes becomes sufficiently large such that (relative to the potentiometer) a significant forward-current occurs, non-linear distortion starts to manifest itself, and the voltage across the potentiometer is subject to limiting. In fact, the output voltage is composed of two parts (potentiometer-voltage, and voltage across the RC two-pole), the consequence being that a part of the undistorted signal is superimposed on the distorted signal. This is a peculiarity of the Tube Screamer (and many similarly constructed devices on the market): it does not only distort but mixes in a bit of the original signal. An easier-to-interpret equivalent circuit is obtained by referencing the output voltage not to ground but to the input connection, and compensating this via adding the input voltage to the output voltage (lower part of the figure).

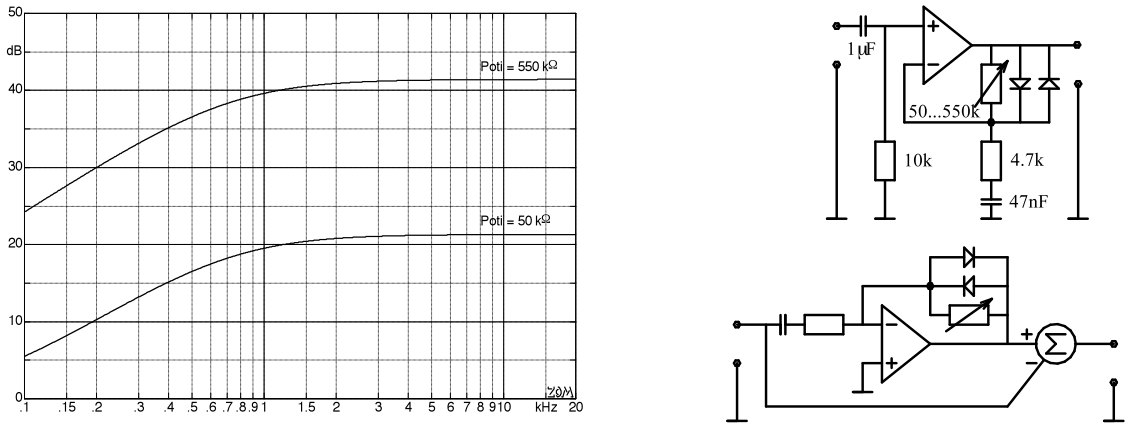


Fig. 10.8.34: Tube-Screamer: small-signal frequency-response and schematic of the distortion section.

Now, the two-part output signal becomes evident: there is the inverted input voltage, plus the (also inverted) high-pass-filtered, amplified and distorted input voltage. Of course, we arrive at the same conclusion using Kirchhoff's loop-rule, and assuming the differential input-voltage of the OP-amp as zero.

The **potentiometer** controls the basic amplification of the distortion branch, but not the amount of the distorted output signal, and not the amount of the undistorted signal, either. The balance between distorted and undistorted output signal is pre-set and cannot be changed without changing the circuit. If, for example, the amount of the distorted signal is to be enlarged, 4 diodes instead of two could be included: two each in series and the two series-circuits in an anti-parallel connection. Using one diode in one direction and two in the other direction creates an asymmetric clipping with a sound that could be considered somewhat fuller and assertive that that of a point-symmetric characteristic. Any preference will be a matter of taste and can – in case of doubt – be changed for very little money. And while we are in the process of changing diodes: a mixture of Ge- and Si-diodes can sound very attractive, and even LED's are deployed these days by Marshall (and everybody else) to achieve distortion.

In **Fig. 10.8.35** we see the time function of the OP-amp output voltages for three different drive-levels. The clearly recognizable phase-shift is not the result of the high-pass at the input but that of the 4.7-kΩ-47-nF-two-pole. The right-hand picture shows a type of distortion that should be avoided at all cost: a piercing-through to the opposite voltage limit. The exact reason (latch-up?) will not be discussed here, but if this happens, another OP-amp-type needs to be brought in. The curves shown here were measured using a TL-072 that – being a FET-OP-amp) apparently is susceptible to such effects. In defense of this actually very good analog IC it should be said that this effect happens only at rather high drive-levels. But if it indeed happens, the sound is so horrible that it probably is usable only as a special effect.

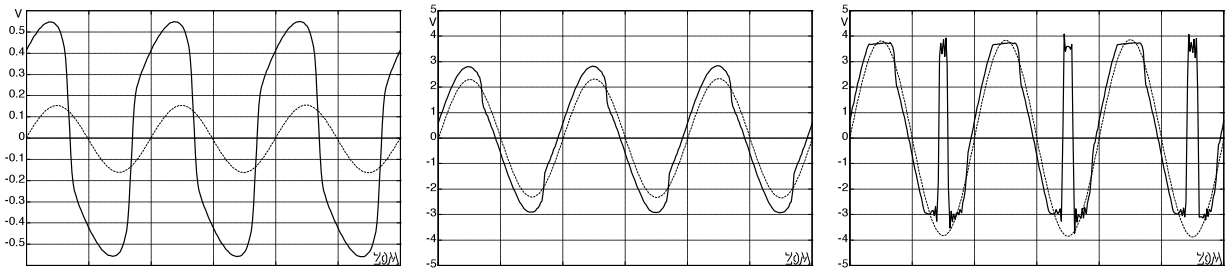


Fig. 10.8.35: Time function for different drive-levels (f = 500 Hz). U_{Batt} = 9V.

In his Gitarre&Bass-article (11/01 – recommended reading!), B.C. Meiser lists several OP-amp types suitable for the operation in the Tube Screamer (e.g. the NE 5532). He also points to the fact that the NJM 4558 is not suitable. A fundamental problem of all pedal-type devices is the requirement that they have to run off small battery-voltages. For the TL-072, the recommended supply voltage is 30 V; in the Tube-Screamer it has to make do with a meager 9 V – and even this only for a fresh battery. The manufacturers do allow for smaller operating voltages, but they do not specify which parameters will then deteriorate. If the specified operating-voltage for the LM1458 is 10…36 V, 9 V is simply too little. The NJM4558 is supposed work from 8 V – but how well will it do the job? In some data sheets we find: use from 12 V. For the Texas RC4558 we read: from 10 V. With regard to the slew-rate, the data given are: for the NJM 4558 = 1V/µs, for the RC 4558 = 0.5 V/µs, and for the MC 4558 = 1.5 V/µs. All these values are specified for 30 V supply voltage and not for 9 V. Trial and error is the only way to find out how well (or how poorly) an OP-amp performs at 9 V; the data sheets give too little information on this. Also, we need to consider that distortion – as it is practiced in the Tube-Screamer – originally was seen as off-limits by the manufacturers. Word has gotten out only rather late that an OP-amp needs to sound good also when overdriven. So: try out some OP-amps – these ICs don't cost a lot.

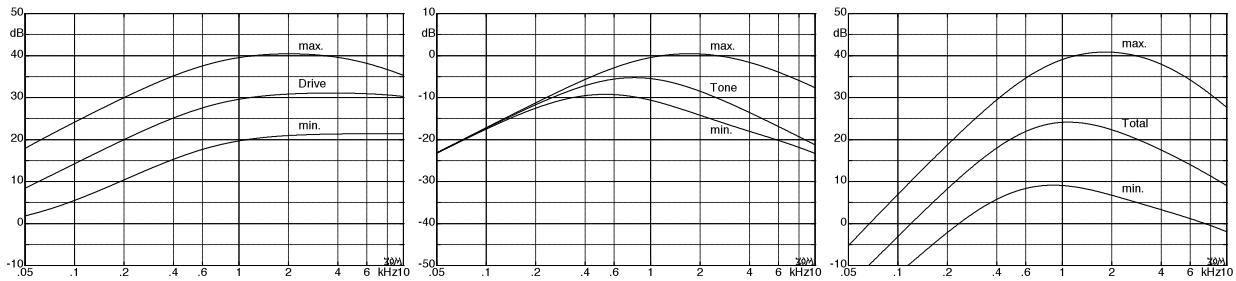


Fig. 10.8.36: Tube-Screamer-frequency responses: distortion unit, sound filter. Overall circuit. The transfer function of the sound filter is easily calculated by Y-delta-transforming the OP-amp input circuit.

10.8.5.5 Fuzz-Face (Dallas Arbiter)

As a typical representative of the group of brute-force distortion-devices, we will in the following analyze the **Fuzz Face**, a small battery-powered effects device offered by Dallas Arbiter♣ from 1966. In the original version, 2 Germanium transistors (AC 128) took care of gain and distortion (**Fig. 10.8.37**); the output voltage could be controlled by the "Volume"-potentiometer while the "Fuzz"-potentiometer adjusted the degree of the distortion.

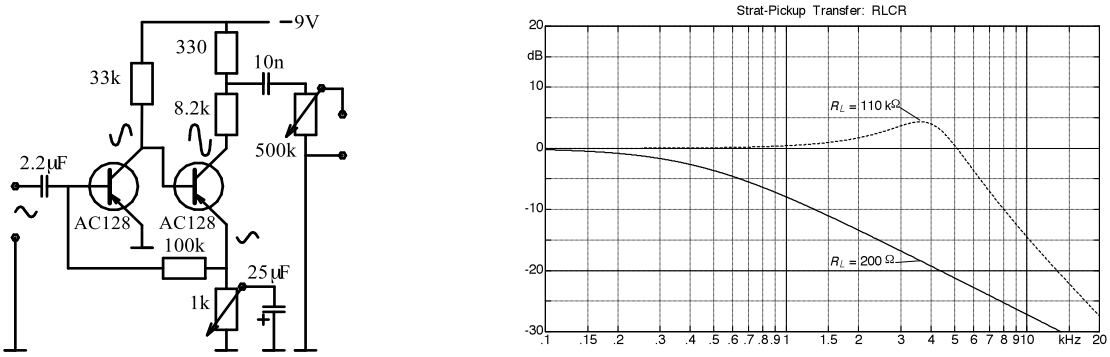


Fig. 10.8.37: Fuzz-Face: circuit (left), pickup-frequency-response (right, compare to Chapter 5.5.4) Details re. the circuit-board construction (and replication) are described by Martin Thewes in Gitarre&Bass, 09/2009.

The circuit is peculiar – starting with the input: due to the current-feedback (100 kΩ), the input shows very low impedance. For the usual pickup, it practically presents itself as a short, especially when the 1-kΩ-potentiometer is set such that the tap is connected to ground. For all measurements described in the following, the generator providing the input signal was connected via a 6.8 kΩ resistor. In this configuration, there is almost zero input voltage – this does not mean, however, that the circuit is not receiving any drive signal. In fact, the input operates under current control as a so-called "zero-ohm-node" known from recording-studio-technology. Due to the frequency-dependent source impedance of a magnetic pickup, the result is a veritable low-pass radically attenuating all treble above 500 Hz. The treble is revived, however, in the form of strong non-linear distortion-products generated via the high gain-factor of 100…2000 (**Fig. 10.8.38**). Because of the current-control, the gain must not be referenced to the input voltage, but to the quotient of collector voltage (T2) and generator voltage ahead of the 6.8-kΩ-resistor. This resistor is required due to the small input impedance; it models the pickup-resistance. Whether the resistor has a value of 5.2 kΩ or 7.3 kΩ is of no importance. As was the case for the Range-Master, charge-reversals in the capacitors cause shifts in the characteristic curve – this to a somewhat lesser extent but with the same tendency.

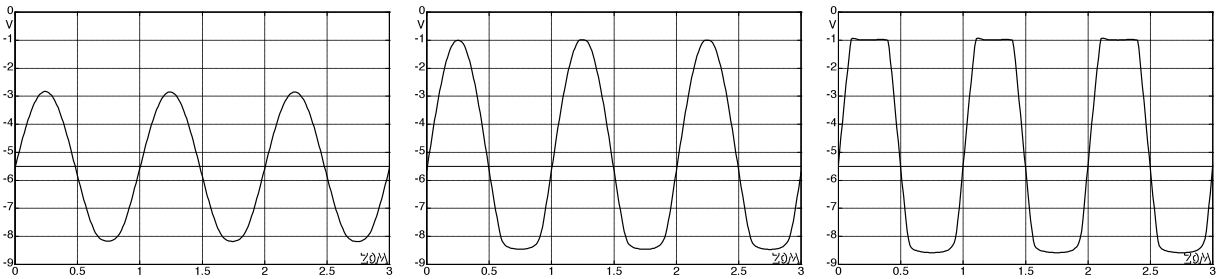


Fig. 10.8.38: Fuzz-Face: collector-voltage of the 2nd transistor for different drive-levels (500 Hz).

♣ First under the "Arbiter Electronics"-moniker, then under "Dallas Arbiter".

Translator's note: *the following page contains a satirical send up of some of the absurdities occurring when transistors and tubes are discussed among guitar "experts". Much of the satire relies on the German language and culture, and it is all but impossible to do a simple translation. Even the title accordingly plays with words: in German, the word (substantive) for "tube" is "Röhre". Incidentally, there is a similarly spelled verb in German: "röhren" that is not related at all in its meaning which is "to roar". The title of this page – literally translated – would thus be: "As the transistor roars". The translated intended pun "As the transistor tubes" does not work, of course, as it does in German where the terms for "tube" and "to roar" both make a kind of sense in the context.*

Therefore this page is left un-translated. Reading it is recommended to all who have a little command of the German language. We might consider translating it at a later point, after all – and even find a native speaker who can come up with correspondences and matches that really ***trans****-late the intended meaning into American or British context and culture.*

10.8.5.6 Wenn der Transistor röhrt

Heute besuchen wir, die Guitar-Licks-und-Tricks-Redaktion, einen Exponenten der deutschen Verstärker-Szene: **Markus Dampfmeister**, den Kölner Amp-Wizard. Mit seinen Marshall-Mods war er schon Anfang der 80er aufgefallen, ist gar bis Straubing gekommen, seither hat sich sein Ruf sogar über die Landesgrenzen hinaus verbreitet. *"Markus, ich darf doch Du sagen?"* "Natürlich, wir sind ja so eine Art Kollegen. Was liegt an?" *"Wir wollten ein altes Thema aufgreifen: Röhren für die Gitarre, oder Transistoren?"* "Röhren, keine Frage. Transistoren geben dem Klang eine kalte Sterilität, deren Dimension man schnell erahnt, wenn man nur mal den Finger auf eine Röhre legt. Dieses Urfeuer, das da im Innern brennt, dieses Elektronenbombardement der Anode, das ist es, was den heißen Ton ausmacht. Mit Transistoren ist das nie erreichbar, die werden, wie der Fachmann sagt, schon bei viel niedrigeren Temperaturen eigenleitend." *"Und dann leiten sie?"* "Nein, das heißt, schon, aber dann sind sie kaputt. Röhren halten da viel mehr aus, die kann man so quälen, dass das Anodenblech glüht. Den damit verbundenen Höllensound bringt der Transistor einfach nicht, einen glühenden 2N3055 hat noch niemand gesehen. Und schon der Fachbegriff: Halbleiter! Gibst Du dich mit halben Sachen zufrieden?" *"Nee, drum sind wir ja hier, um endlich einmal die volle Packung zu bekommen. Also Röhre?"* "Nur! Die alten sind die besten! >Hast Du Tungsol in dem Fender, gibt's nur eins: Return to Sender. Dieses Mumpfen, dieses Dröhnen – schauerlich, zum Abgewöhnen. Ist es aber eine Mullard, kriegst du einen Sound, der pullert<. Ich habe vor Jahren alte UK-Bestände aufgekauft, die ich jetzt gegen echtes Geld in meine Customs einbaue." *"Und Röhren haben keine Nachteile?"* "Nee. Sie rauschen sogar weniger als die Transistoren. Weil: Die ganzen zwangsläufigen Verschmutzungen, die im Silizium (Sand!) unvermeidlich sind, die hat die Röhre nicht. Sogar die Luft wird rausgesaugt, damit ja nichts stört. Und nochmals: Die alten sind die besten – damals war einfach auch die Umwelt noch nicht so versaut, da konnten sie noch hochwertige Materialien verbauen. Ergo: Sound pur."

Der Mann wusste, wovon er redete, das war noch einer der alten Schule. Thorben, unser Fotograf, hatte bisher Pics von der Location geschossen, nun meldete er sich plötzlich mit einer Frage: *"In den Verzerrern sitzen aber Transen, oder?"* "Ja, drum rauschen die auch so stark. Da muss man aber höllisch aufpassen, da gibt's Si und Ge. Ge-Transistoren klingen gut, Si kratzt und sägt. Bei Si ist nämlich der sog. Bandabstand größer, da tun sich die Elektronen schwerer um rüberzuhüpfen. Mit einem alten OC44, der schon mal für 10 Euro gehandelt wird, lebt jeder Verzerrer auf, da entstehen einzigartige Klangwelten, das lebt, das röhrt, wenn man so sagen will." *"Aber vor einem Monat waren wir beim BCM, und der meinte, anstelle eines OC44 kann man locker einen OC71 reinstecken. Oder einen OC75."* "Ja, das sind alles alte Germanen, da ist praktisch dasselbe drin, nur mit anderer Aufschrift." *"Oder einen AC122, oder einen AC128..."* "Ja gut, was der BCM wo reinsteckt will ich jetzt nicht kommentieren, wenn das der BCM meint ... das sind europäische Vergleichstypen." *"Oder einen OC76, oder einen OC77. Ganz gut soll der amerikanische 2N508 sein, sagt der BCM, oder der 2N527 ... ich versteh die Sache mit der Einzigartigkeit noch nicht so ganz."* Thorben, der Schrecken der Redaktion, immer gut für einen Eklat. Markus verlor zusehends die Lust, über die Problematik ubiquitärer Einzigartigkeit nachzudenken, er wollte schließlich seinen neuen Custom in diesem redaktionellen Beitrag unterbringen, in Zeiten wie diesen ist man um jede kostenlose Werbung froh. Doch an Thorben führte kein Weg vorbei: *"Ich hab noch nicht verstanden, warum einerseits die glühende Röhre für zerrende Höllensounds am besten sein soll, andererseits der Clapton* (und nicht nur der, d. Red.) *vor seinen Röhren-Amp einen Range-Master gesteckt hat, in dem doch, wenn ich mich nicht täusche, ein Transistor sitzt?"* Das fasste Markus nun allerdings als unfreundlichen Akt auf, der Talk endete unplanmäßig. Schade eigentlich, ich hätte noch Fragen gehabt. Macht aber nix, nächsten Monat geht's zum Piepenbrink, von diesen Exoten hat's ja noch mehr.

10.9 General operating characteristic

The topic of the previous chapters was the performance of individual tube stages or individual circuit sections; in the following the focus shall be shifted to higher-level considerations. The operating characteristic of a guitar amplifier can be looked at from two different angles: from the point of view of the circuit design (i.e. how does the circuit function?), or from the point of view of auditory acoustics (i.e. how does the amp sound?). Of particular interest is to causally interconnect the two approaches – that, however, is also the most difficult task.

10.9.1 Tube sound vs. transistor sound

In fact, transistors seem to have only advantages over tubes: they are smaller, cheaper, have no fragile glass containers, do not need heating. Apparently, they have the single disadvantage that guitar amps designed using them do not sound good. Of course, this is a highly subjective judgment, and of course there are other opinions – however, in particular the early transistor amps had few advocates, and aside from all mysticism there are without a doubt differences from the point of view of systems theory. Still, there is no "tube sound" *as such*, just as there is no "transistor sound" *as such*. A guitar amplifier does not sound better merely because it is fitted with tubes, and a transistor amp does not need to inherently sound bad. It might, though. Fender's Solid-State-Series – the one advertised in 1968 with *'superb sound'* letting you *skim the waters of musical greatness* – was rather unsuccessful. *'Curious refrigerators'* was the term used by the German Gitarre&Bass-magazine in their Fender special edition. To vindicate the name "Fender", one could argue that "this wasn't Fender, this was CBS"; however, after the sale of his company to CBS, the same Leo Fender designed and produced (with his new company Music-Man) hybrid amplifiers featuring a transistor pre-amp and a tube power-amp. These amps – at least today – by far fail to achieve the fame and glory of Fender's black-faced heroes. The same with VOX: the company did not become famous with the transistorized Defiant, but with the all-tube AC30. Guitarists and transistor amps: not a love at first sight.

Transistor amps sound sterile, impersonal, lifeless, they buzz, crackle, sound scratchy, and on top of everything, at the same wattage they are not as loud as tube amps. These subjective judgments elude any circuit analysis. Who wants to stipulate how a guitarist should perceive his guitar sound? Plus: even if this is pure imagination, it is easily conceivable that this kind of imagination has repercussions on the virtuosity. Electrical engineering with its many disciplines is actually only challenged when causal links are brought in: *'hot tubes for a warmer sound'*, or: *'tubes do a rounder limiting and thus sounds less sharp'*, or: *'tube amps sound better because 2nd-order-distortion dominates in them'*. Still, it is not that simple. If the audible differences in sound could be traced to a single reason, we would probably see exclusively transistor amps today. As is the case for public address systems: who would make the effort to stage several hundred tube power amps? But nine tube amps lined-up behind a guitar player: even today, that is not very strange. "Very loud" in the case of nine AC30, and "VERY LOUD" if six Super-Twins are stacked to a pyramid. Why do they do that – what is the secret of the tube? With such a presumptuous question, the answer can only end up in hybris … anyway: the secret i.e. the undiscovered country is in the diversity, in the interaction of a multitude of non-trivial components and characteristics, respectively.

Harmonic distortion, slew-rate, frequency response, input- and output-impedance, shifts of operating points … and all that in combination. What is the effect of the level-dependency of the 4th-order harmonic-distortion on the sound? Does the 4th-order distortion have to be even considered at all, and if yes, up to which order are distortions relevant? To measure the distortion is simple, but to determine its effect on the sound is difficult. For one, comprehensive auditory tests are required, and then each judgment is dependent on many boundary conditions: on the setting of the tone controls, on the loudspeaker, on the listening location, on the guitar, and of course on the generated tones. Because this variety of parameters is vast, the developers of transistor amplifiers do not only have to develop circuits but also "survival strategies".

One of these strategies is: *as long as the frequency response is identical, the sound has to correspond, as well*. That thought is too simple. Here's another: *since it is unknown how the characteristics of each individual tube stage affects the sound, every detail of the tube circuit needs to be modeled*. Sounds as if it would be on the safe side – and it would be if indeed every detail were known. And another variant: *we tune and retune until everybody is satisfied, even if the new circuit has no relations to anymore to tube circuits*. Possible, but easily subject to diffuse criticism: something is missing! No one knows exactly what it is that's missing, but everybody is convinced: that is not the ideal tube-sound. Or the opposite approach: the designers are happy (sounds quite good and doesn't go up in smoke anymore), the management as well (remained even 2% below the pre-calculation), and the sales department agrees (finally they finished it). It's juts that not only the calculation but also the turnover is below plan. It is a difficult market: the original Bassman is a legend but the Music-Man is not. Despite the fact that behind both the mastermind was Leo Fender.

This book with its focus more on guitars will not answer the question which type of distortion will render the sound sparkling-creamy-wooden-throaty – that topic belongs to a book exclusively dedicated to amplifiers. Still, there is room for a few basic thoughts. The previous chapters dealt with the non-linear behavior of the amplifier; from the point of view of the author, this is a main theme. In tube amplifiers, several linear and non-linear systems interact: *high-passes* in the coupling-capacitors, in the output transformer and in the loudspeaker, *low-passes* in every tube, in the output transformer, in the loudspeaker. In every tube, in the output transformer, and in the loudspeaker we also find non-linearity. It all makes for an almost unfathomable system – even without negative feedback. It's not that we couldn't describe the individual sections of the system – it is the overall judgment that is so difficult. Something that is routine in the LTI-system develops into a vast problem for coupled non-linear systems. For example, it is only in the linear system that it makes no difference whether filter-poles are realized in the electrical of the mechanical domain – here we can compensate e.g. a treble-loss of a loudspeaker by an electrical filter. If indeed linear behavior is desired, equal-sounding amplifiers can easily be built with both tubes and transistors. With non-linearity entering the picture it gets very complicated, though.

After several decades of searching for the right sound, transistorized guitar amps today have matured to the point where the acceptance can to said to be good. Nevertheless there are still innumerable tube amps on the market, and many guitarists are likely not to buy anything else for decades to come, ready to invest those 200 € now and again in a quartet of tubes. The manufacturers have learned from the mistakes of the early years, and offer well-sounding circuits and amps. However, the guitarists have also educated themselves and now can hear details that 50 years ago would have been classified as unsubstantial.

What adds to the problems that non-linear amplifier circuits can pose are **psychometric** issues: how do we measure auditory perceptions? Here, the range starts with "plug in, turn up, listen" and ends with round-robin-tests carried out on a global scale. For guitar amps, we mostly find the former experimental method: listening test in the store, in the rehearsal room, in the editorial office. The results of such test are often ignored or contested on the side of the acousticians (who typically have a scientific background) – not so much because of the involved jurors (normally not that wee known) but because of the unscientific approach lacking objectivity and reproducibility. Is the hand-wired boutique-amp praised primarily because it hails from California and sets you back 5000.- €, as every person testing the amp is made to know first of all? Would the amp be as convincing if it would remain unknown and hidden behind a curtain? We find a nice example for such a scenario described by Uli Emskötter in SOUND-CHECK magazine [issue May, 2000]: in a test of guitar cables, all involved perceive "*pronounced differences in sound*". A week later, in a repetition – this time as a blind-test – shows: "*bewildering result – the judgment came out as entirely different*". It is nothing new to psychologists that the type of presentation procedure will influence the result of perceptional tests. The insights these experts have regarding **experimental methodology** are beneficial for listening experiments, as well (see Chapter 10.9.4).

Guitarists not only perceive the sound of their guitar but they also evaluate it. For the **perception process**, a relatively small inter-individual variance may be assumed, however the **evaluation process** always depends on various boundary conditions. We all know this: first the new CD by the latest superstar is lauded to be among the 10 best releases of the year, next thing we know it appears in a TV-show listed among the most embarrassing oeuvres. Although the auditory event remains exactly the same, the evaluation of it changes. Another example that every studio musician is familiar with: you do a mix-down, find a suitable mixer-setting, everybody is delighted and calls it a day. The next day you listen again – without any changes to any setting – and everybody is disappointed: the vocals are too loud, the drums are trebly, the bass too fat … or the other way round. The reasons for this change are rarely found the technical issues (loudspeakers cooled down, humidity different) but with all likelihood the difference is found in the changed judgment-standard. The underlying processes may develop over minutes or even hours, but time-invariant, systematic differences (bias, offset) are also known: there is a tendency to set a value controlled by the subject to high [12, loudness scaling].

Here is an episode showing how much our value judgments are affected by cognitive processes. It may be a singular case, but is likely to happen quite often in a similar way. After a gig, a young musician addressed me speaking in highest terms of the *"super-sound"* my guitar-rig had: *"You can't beat the good old AC-30 – that's pure tube-sound."* I am sure the people at VOX would have loved to hear that – although I wasn't using an AC-30. At the beginning of the 21st century, this legendary amp is not as ubiquitous as it used to be, and the younger generation apparently is not as familiar with it. Indeed it was a VOX I was playing through, as the gilded badge on the front of the amp confirms, however it rather was a AD-60-VT. That amp features a transistor-preamp and a transistor-supported 1-W-tube-output-amp. It doesn't sound that horrible, either, in fact it sounds pretty darn good, and its **AC-30TB** model cooperates most harmoniously with the Historic Les Paul. In any case, associating the terms *VOX = AC-30 = tube amp = super sound* appears to be hard-wired into many (though not all) a musician's brain. Had the amp-label not read VOX but Solid-State-MOSFET, the judgment could easily have been "*doesn't sound too bad – for a transistor amp*". Indeed the psyche plays an important partner-role in the wide and colorful world of psychophysics. The psyche's counterpart in this area, i.e. physics, and more specifically circuit technology, will now get some attention as well.

From the many amplifier circuits we chose a Fender- and a Music-Man-circuit (**Fig. 10.9.1**) because Leo Fender was involved both. He probably was not the only designer but at the very least the responsible patriarch. Starting out from the RCA-application-notes, the circuit of the **Twin-Reverb** – as it is presented here – was developed over the years into an internationally recognized standard that inspired competitors, as well. Once the era of the octal-tubes had passed, noval-tubes entered service at Fender in the mid-1950's, and in particular the high-gain 7025 (and its colleagues, the 12AX7 and the ECC83) won the pole-position that they never relinquished again.

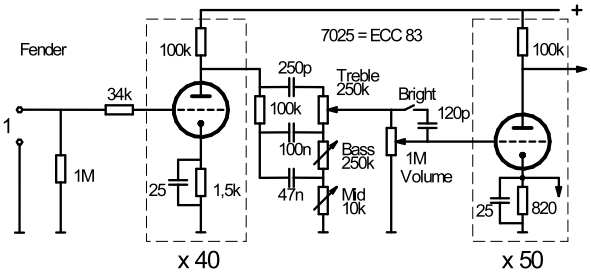


Fig. 10.9.1a: Fender AA763 (Twin-Reverb).

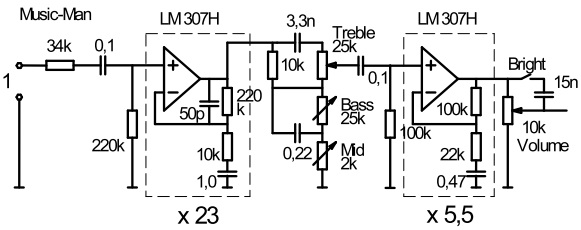


Fig. 10.9.1b: Music-Man 2100

The Fender-circuit starts with a triode-input-stage of a gain factor of about 40. Tone-filter, volume control and intermediate amplifier follow. The Music-Man circuit shows considerable similarity although the input OP-amp has only a gain factor of 23 – a concession to its lower supply voltage that leads to earlier limiting. Despite small differences, the effects of the tone-filters are comparable. The volume pot, however, is connected not directly after the filter but inserted after the intermediate amp. First, these circuits will be compared regarding their linear behavior.

Differences show up already at the input: the MM has lower impedance than the TR: the pickup-resonance receives a stronger dampening. On the other hand, the input capacity is lower in the MM (do consider the Miller effect!). The series-capacitance in the MM-input has barely any effect on the signal, and no shift of the operating point need to be feared, either (due to the symmetrical limiting in the OP-amp input). The 50-pF-capacitor is there to reduce the gain at high frequencies, it has an effect from about 10 kHz The 1-µF-cap reduces the gain at very low frequencies (below 20 Hz). Compared to the TR, the impedance level in the MM-tone-filter is lowered by a factor of 10 to normal OP-amp-typical values. Disregarding this change, the two tone-filters are indeed very similar, despite one missing capacitor in the MM – however such variants with only two capacitors did exist at Fender, as well (e.g. the Super-Amp, see Chapter 10.3). These circuits were modified again and again.

The small-signal transmission factors of both circuits are shown in **Fig. 10.9.2** (referenced to 1 kHz for both). This similarity is not likely to have been an accident; rather the Fender circuit will have been the given objective. The only significant difference in the small-signal behavior is the different input impedance; we can only surmise that the design process was possibly checked with a low-impedance generator such that this aspect did not become apparent. Large differences are apparent, however in the behavior for strong signals i.e. at high drive-levels.

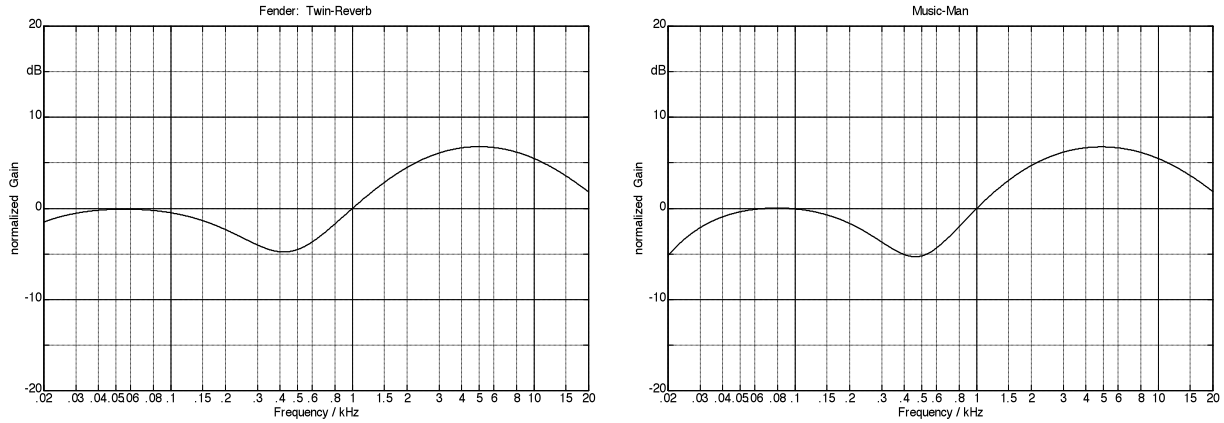


Fig. 10.9.2: Frequency responses referenced to 1 kHz. Bright-switch in "off"-position for both cases.

The drive limit of the amplifying element (tube, OP-amp) is the significant factor for the behavior at high drive-levels. Differences pop-up right at the input: in the TR we have grid-current distortion that is not there in the MM. From about 5 kHz, the MM shows slew-rate distortion but the TR does not. Highly significant: the TR has the volume control positioned after the first tube while the MM has it only after the second OP-amp. With the treble-control turned up fully, the gain factor in the MM from input of the first OP-amp to output of the second OP-amp is about 126. To maintain distortion-free operation, the input voltage must not increase beyond 70 mV. For the TR, this is quite different: assuming 35 V as limit of the first tube for hard clipping, the permissible input voltage would be about 900 mV. Having said that: as already mentioned in Chapter 10.1.4, it is difficult to compare tube- and OP-amp-distortion. Below the clipping-limit, the OP-amp works practically distortion-free, while for a tube, distortion rises continuously across the drive-level-range. **Fig. 10.9.3** shows, for the MM, the maximum input level for undistorted operation. Especially in the brilliance-range (3 – 5 kHz) that is so important for Fender guitars, distortion can very easily occur even though the volume pot may turned up only a bit. In Fender amps, the **Bright-switch** most often is located at the volume control, but for some MM-amps this is included into the negative feedback of the first OP-amp – possibly to reduce OP-amp-noise. Switching-on the Bright switch in the latter case further decreases the treble headroom (right picture), independently of the position of all tone pots and of the volume control. This marks a difference to the Twin-Reverb and to similar Fender amps. The MM-amps therefore do show clear differences in their behavior at high drive-levels compared to typical tube amps.

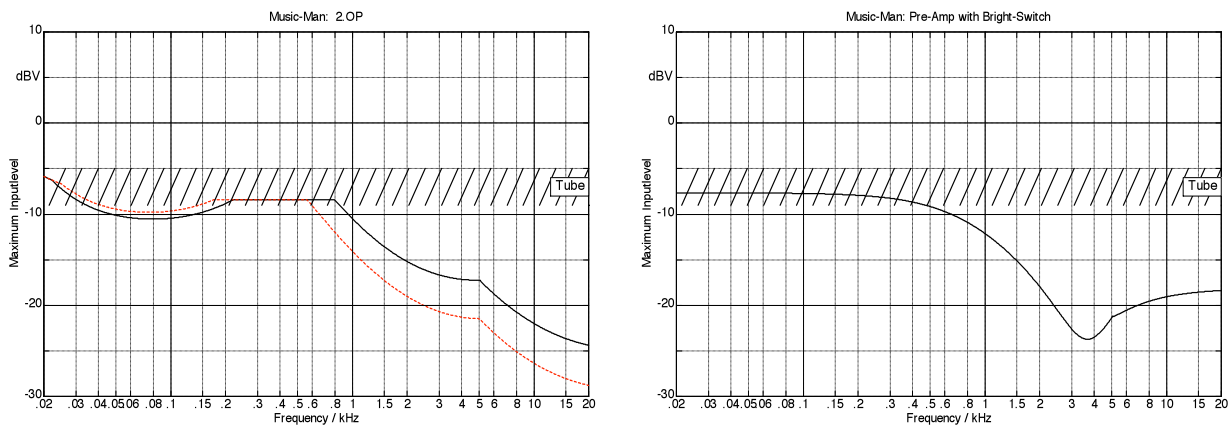


Fig. 10.9.3: Music-Man: maximum input level for undistorted operation. **Left:** solid line = tone controls as in Fig. 10.9.2, dashed line = treble control turned up fully. **Right:** amplifier in which the Bright-switch changes the gain of the input-OP-amp. Hatched area: tube input-stage.

10.9.2 Tube-Watt vs. Transistor-Watt

Allegedly, tube amps are louder than transistor amps rated at the same output power. Before we discuss this topic on a scientific level, we first need to establish what exactly is meant with "this is a 50-W-amp". *Not* meant is the power consumption i.e. the power drawn from the mains. Rather, such a specification always refers to the output power fed to the loudspeaker. If the speaker impedance were frequency-independent and real, this power could be stated without any issue. However, the loudspeaker impedance is frequency-dependent and complex, despite the simple 8-Ω-label. In order to still be able to specify a number of watts, an ohmic resistor replaces the speaker, and it is for this resistor that the indicated number of watts holds. In other words, the manufacturer specifies that an amp generates e.g. 50 W at 8 Ω. This does not in fact tell us how much power this amplifier can deliver into an 8-Ω-loudpseaker, because an 8-Ω-speaker does not have 8 Ω at all frequencies (Chapter 11.2).

In an 8-Ω-resistor, an alternating current of the RMS-value of I = 2 A generates the RMS-voltage U = 16 V; the product of current and voltage yields the power: P = 32 W. In order to clarify that these are RMS-values, a tilde is often put over the formula symbols:

$$P = \tilde{U} \cdot \tilde{I} = R \cdot \tilde{I}^2 = \tilde{U}^2 / R\,; \qquad \tilde{U} = \sqrt{P \cdot R}\,; \qquad \tilde{I} = \sqrt{P / R}$$

For an RMS-current of 2 A, the matching current amplitude is $\sqrt{2} \cdot 2$ A (i.e. 2,8 A), and correspondingly the voltage amplitude of 22,6 V matches an RMS-voltage of 16 V. Multiplying the amplitude-values rather than the RMS-values yields double the power: a 32-W-amp turns into a 64-W-amp. This (higher) wattage specification is not in use in the professional audio technology – rather, the nominal power calculated from the RMS values is specified; in the example this is P = 32 W. What does this power depend on? Its factors are e.g. the squared RMS-voltage, and a more or less arbitrarily defined nominal resistor R that initially replaces the loudspeaker. The resistor is defined as fixed quantity in the data sheet; the voltage is, however, variable. So, for which drive-levels do we specify the nominal power? For studio- or HiFi-equipment, the largest voltage just below distortion level is used, or the voltage at which a certain total harmonic distortion (THD, to be specified) occurs: e.g. 32 W at 8 Ω and for k = 1%. For a guitar amplifier, such a THD-specification is not possible, and therefore – for a sine-shaped drive signal – the output voltage is visually judged to specify at which level clipping occurs. Again: for the calculation this limiting voltage may not be substituted into the formula because it represents the amplitude (i.e. the peak voltage). Rather, this limiting voltage needs to be divided by $\sqrt{2}$. Alternatively, the amplitude is used, and the calculated power is then divided by 2. As an example: for an 8-Ω-resistor, clipping occurs at 40 V. The resulting RMS-voltage is 28,3 V, and the power is calculated to P = 100 W. Alternatively: $40^2 / 8 / 2 = 100$.

Incidentally, it is not sufficient that a loudspeaker box connected to a 100-W-amplifier can withstand merely 100 W. Since guitar amps are typically overdriven, they generate more than the specified nominal power. Given that the nominal power mentioned in the above example is independent of the load, for a square-shaped signal the power would be double, i.e. 200 W! This is because square- and sine-shaped signals of the same peak-value differ by a factor of $\sqrt{2}$ in their RMS-values.

The limiting-voltage (i.e. the voltage at which the output voltage starts to clip) is, however, not entirely independent of the load because the internal impedance of the power supply is not zero. The **power supply** furnishes the operating voltage to the power amplifier – for a tube amp e.g. 450 V. This is a dc voltage the value of which depends on several parameters: on the mains voltage, on the power transformer, on the rectifier, and on the load. In the unloaded state, the operating voltage has its maximum value but buckles ("sags") under load, i.e. as the amplifier feeds power to the loudspeaker. This has a very simple reason: the current flowing to the power amp first needs to pass through the mains-transformer and the rectifier – and either of them causes a voltage drop. The exact voltage and current time-curves are anything but simple to describe (these are coupled non-linear systems), but we do not need to examine this very precisely here. With a load connected, the operating voltage buckles and decreases, e.g. from 450 V down to 400 V, or even down to as low as 360 V. Given a large mains-transformer and an efficient silicon-rectifier, the voltage drops only little; with a small transformer and a tube-rectifier the drop is larger – this is another genre-typical difference. Massive 100-µF-capacitors make the "sagging" (as well as the subsequent recovery) slower than the (from today's perspective) puny little 16-µF-caps. Here we actually may have a difference between tube-Watts and transistor-Watts: modern transistor amps often have very "stiff" power supplies, i.e. power supplies with a small internal impedance the voltage of which decreases only little as a load is connected. Tube amps (especially if they are from back in the day and carry a tube rectifier) have power supplies with comparatively larger internal impedance (see Chapter 10.1.6). Of course the two aspects are not necessarily connected to each other: a tube power amp could just as well include a power supply with low internal impedance – but in particular the legendary amps do not. For a guitar-note played after pause, the full charge of the power-supply-cap is available during the first instant. The limiting voltage may e.g. be 40 V yielding 100 W of nominal power into 8 Ω. However, the voltage buckles after a few milliseconds and the limiting voltage drops to e.g. 35 V. With the power being in a square-dependency to the voltage, the power decreases to 77 W. Measuring the nominal power with a continuous sine-tone yields the second value, i.e. 77 W. For a transistor amp fitted with a "stiff" power supply, the limiting voltage may decrease e.g. only from 37 V to 35 V, so that both amps have the same nominal power. For an impulse, i.e. as a string is struck, the tube amp does however have a higher power; in the example it is 100 W rather than 85 W. In case the limiting voltage of a tube amplifier does not only decrease by 12,5% but by 15% or 20%, these differences become substantially larger.

Thus, one difference in the power yielded by tube- and transistor-amps relates to the temporal behavior: the "attack" is delivered with more power in a tube amp. This holds for the generic circuits – of course it could be designed exactly the other way round. Consequently, the theorist is of course right as he states: "there is no difference between tube-watts and transistor-watts; watt as the unit for power is universally standardized". However, in just the same way the musician is correct in perceiving his or her tube amp as louder. It is not the unit of measurement that is different but the measurement process. A second difference is found in the resistance of the loudspeaker that is not constant, but frequency-dependent and complex. The magnitude of this complex resistance, the **impedance**, may easily reach 20 Ω or 30 Ω at certain frequencies although the loudspeaker is specified at 8 Ω. Not only the copper-resistance of the voice coil contributes to the impedance but also the inductance of the voice coil and the moving mechanic component as they are transformed into the electrical domain (Chapter 11). At the resonance frequency, the loudspeaker assumes high impedance, and the same happens at high frequencies.

Fig. 10.9.4 shows the frequency responses of some guitar speaker boxes: the changes with frequency are most obvious. It is rather up to the manufacturer which impedance value he specifies for the respective box. There are indeed standards for this, however the musician and the manufacturer do no actually shake hands over a sales-deal based on specific DIN- or ANSI-norms. For the following consideration, we simply assume the loudspeaker impedance to be 8 Ω at *one* frequency, and 16 Ω at *another* frequency. If the amplifier has a transistor-typical "stiff" power supply and features an also transistor-typical strong negative feedback, the output voltage will be impressed i.e. almost independent of the load. With a 16-Ω-load, the amp will merely be able to feed half the power that it can generate in an 8-Ω-load. The situation is very different for a tube amplifier: operating it without a speaker could even cause flashover at the power tubes – the voltages that may occur are that high. The tube amp is not actually a true current source, but it does feature higher internal impedance compared to a transistor amp. This has consequences on the power delivery. For example: an amplifier with 8 Ω internal impedance feeds P_1 = 50 W into 8 Ω and P_2 = 44 W into 16 Ω. A (transistor-) amp with 0 Ω internal impedance would generate 50 W and 25 W, respectively. As the loudspeaker impedance increases, the power delivered by a transistor amp will decrease more strongly than for a tube amp. Again, the exact calculation is rather complicated because linear behavior (internal impedance) and non-linear behavior (limiting voltage) interact, and also because not a sine-tone but a guitar-signal drives the amp. Still, the statement remains: your typical tube amplifier will generate on average more power into a loudspeaker than a transistor amp having the same nominal power rating.

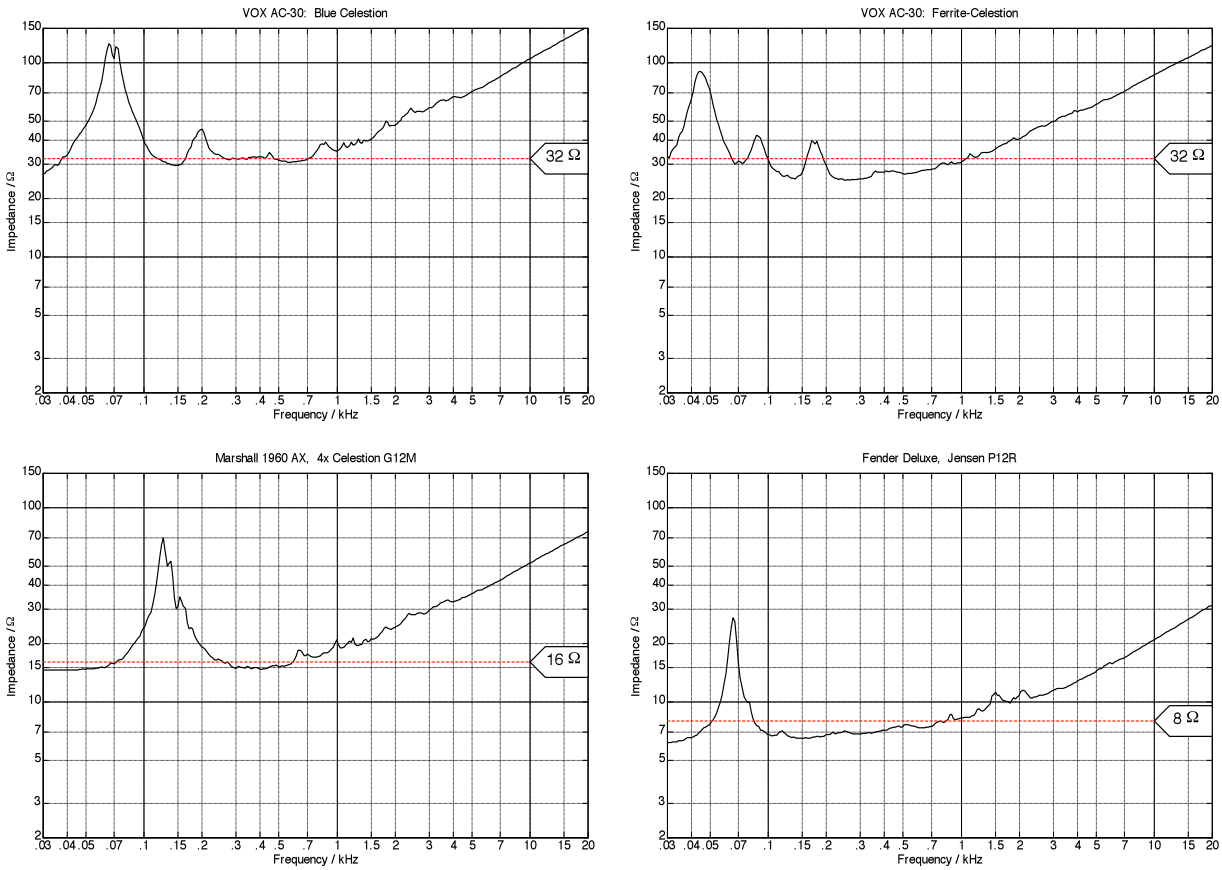


Fig. 10.9.4: Frequency responses (impedance) of typical guitar speakers; measured in a reflecting environment.

As an example we will look more closely at the frequency response of the speaker impedance of a Marshall 1960 AX speaker. It is specified at 16 Ω, its minimum impedance is 15 Ω.

Z reaches its maximum (70 Ω) at 130 Hz. A transistor amplifier rated for 16 Ω and fitted with a (ideally) stiff power supply will feed into 70 Ω merely 23% of the power that it could feed into 16 Ω. In reality, the power reduction will not be as pronounced because the supply voltage will sag less at increasing load-impedance – a reduction to "only" 30% is nevertheless quite drastic. A tube amplifier will behave quite differently: if it is specific for operation with 16-Ω-load, as well, we could expect 60% of the power at a 70-Ω-load, after all – double of what the transistor amp could generate. With a tube amp, the Marshall box will emphasize the frequency range around the resonance frequency, and it will reproduce the treble range more strongly. This tendency cannot be compensated for in the transistor amp by increasing the gain at high frequencies (e.g. by turning up the treble control) because that measure does not influence the maximum power delivery.

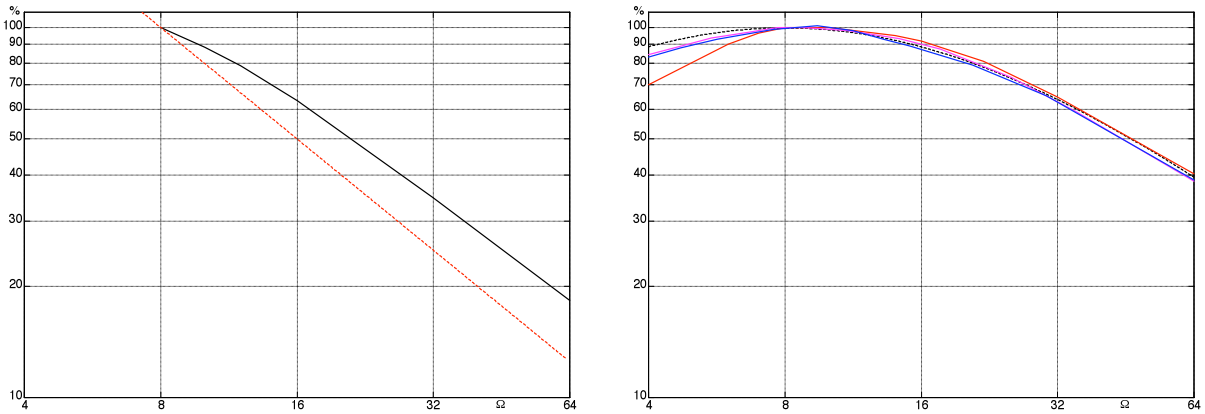


Fig. 10.9.5: Maximum available power dependent on the (ohmic) load resistance. Nominal impedance: 8 Ω. Left: typical transistor power-amp. Right: typical tube power-amp. Dashed: model-calculation.

In **Fig. 10.9.5**, the maximum available power is shown for a typical transistor power-amp and for three tube power-amps, respectively. "Maximum power" means total overdrive. The transistor amp is specified for 8 Ω; for lower loads the amp shuts down. The tube amps are also specified for 8 Ω but can deal with lower as well as with higher load impedances. For the transistor amp, a load-independent imprinted voltage was used as idealized **model**, while for the tube amps a constant internal **impedance** of R_i = 8 Ω was assumed. When discussing the internal impedance of a power amplifier, we need to distinguish between linear and non-linear behavior: during linear operation (no overdrive), the typical transistor amplifier features a very small internal impedance (e.g. 0,1 Ω or even less), while a tube power amp without any negative feedback (such as the VOX AC30) possesses e.g. 80 Ω (there are several variants). The AC30 therefore emphasizes already in its linear operational mode those frequency ranges where the loudspeaker features high impedance. In non-linear operation, the internal impedance can only be defined using special model laws; the dashed line in Fig. 10.9.5 was calculated for tube amplifiers and R_i = 8 Ω. Again, the frequencies of higher speaker impedance are emphasized although not as much.

The **VOX AD60-VT** realizes an interesting concept: this guitar amp uses a weak double triode (ECC83) as push-pull power amp and supplements the missing power via transistor-support. The peculiarity here is that the speaker impedance influences the power that the power amp is able to muster. The lone tube is not included as an alibi, as both power-measurement and listening tests prove (Fig. 10.9.6). What is not advertised as loudly, is that in the AC-30-power-amp, pentodes (EL84) do the work while in the AD-50 VT, triodes are on the job. They do, however, this job with very good success.

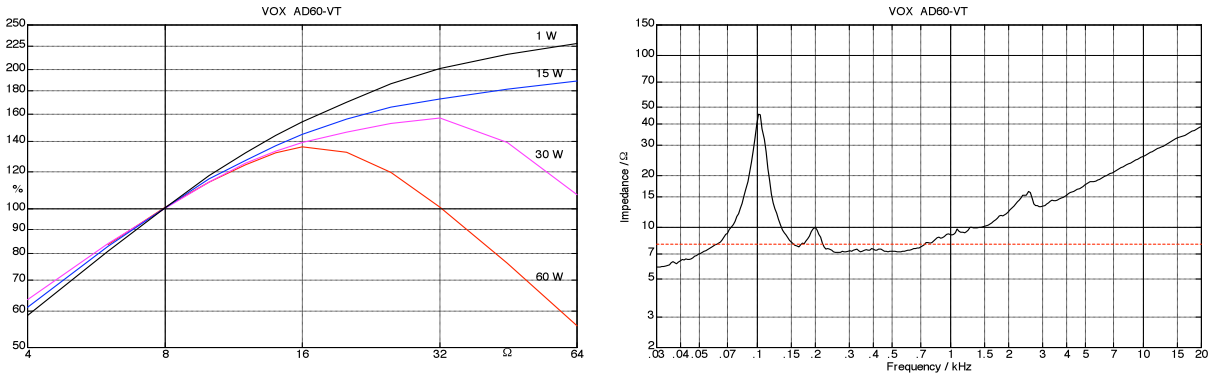


Fig. 10.9.6: VOX AD60-VT: standardized maximum power (left), frequency response of the impedance (right).

Fig. 10.9.6 depicts the maximum offered power dependent on the load resistor. The power-range selectable via a switch is the parameter. The power-characteristic is not really identical to that of a tube amp but the result is quite easy on the ears. Relative to the 8-Ω-reference, a boost can be seen and heard in the frequency ranges with higher speaker impedance – this boost is even stronger than that found with a "true" tube amp. In the 60-W-mode, the power maximum is at 16 Ω – this supports an operation with a (serially connected) second 8-Ω-speaker: more power is available although the treble boost effect is now getting a raw deal.

To summarize: both the impulse power (also termed peak power) and the power delivered in the higher-impedance frequency ranges of the speaker is higher for a typical tube amplifier than for a typical transistor amplifier – with are both rated at the same nominal power for the same nominal load resistance. A percentage value of the difference that would be generally valid can, however, not be given, since the individual circuit concepts are too different.

In closing we should quickly also visit the issue of **loudness** – which in the end is the main aspect of interest to the musician. It is well known that doubling the amplifier power will not always double the loudness. On the other hand, the rule taught in psychoacoustics that for doubling the loudness the 10-fold amplifier power is required, only holds for a 1-kHz-tone. The guitar generates a broadband sound that does not share much with a pure tone, and this naturally needs to be considered. However, of even more practical importance is the fact that the musician judges the loudness of his or her instrument based on how well it can (sonically) hold its own relative to other instruments. In this scenario, the absolute loudness is not as important as the so-called **partial masked loudness** [12]. For example, we may think of a keyboard player sounding a loud chord, and of a guitar player remaining unheard[1] although his amp generates 10 W into the loudspeaker. The latter is not broken at all, but the sound it radiates is fully masked by the sound of the keyboard. As the power of the guitar is increased (e.g. to 20 W), the guitar becomes audible. However, as long as the keyboard is sounded, the loudness of the guitar remains a partial masked loudness and the guitar will be perceived softer compared the loudness it would have when played on its own. For the increase of the partial masked loudness the simple 10-dB-per-doubling-of-loudness law does not hold; a smaller dB-value is valid, e.g. merely 3-dB per loudness doubling. That way, relatively small power-differences gain a bit more significance than basic psycho-acoustical know-how would acknowledge. We shouldn't overdo it, though. The difference between a 50-W-amp and a 55-W-amp remains insignificant. The exact location of the perception-threshold can only be established for each case individually because the masking effects are dependent on the temporal and spectral structure of the involved sounds.

[1] *Translator's comment: the guitar not loud enough - as if that ever happened! Not a realistic example, it seems. Maybe the other way 'round ..... the Leslie for the Hammond won't ever match the Marshall stack, anyway ...*

10.9.3 Coupling capacitors

Coupling capacitors? For some just a cheap run-of-the-mill product about as portentous as a tapping screw, for others a true object of desire worth investing the occasional 50 €. The coupling capacitor separates the AC-part of the plate voltage from the DC-part – it "couples" the AC-signal to the next stage in the circuit. A plate voltage varying between e.g. 200 V and 300 V may alternatively be seen as 250 V DC voltage with a superimposed AC voltage (of an amplitude of 50 V) – the coupling capacitor passes the ac component and blocks the dc component. Let's ignore, to start with, that the coupling cap in fact doesn't let anything through because there's an insulator inside, and let's also disregard that this insulator, on closer inspection, does not perfectly insulate, either. A capacitor lets through AC and blocks DC – that is a good first working hypothesis. We may – or may need to – modify it when necessary. If it is that simple, why do "those in the know", the self-proclaimed amp-gurus, report with a conspiratorial vibe that the long-desired sound presented itself only after swapping the coupling caps? And why is it that – depending on which camp one seeks association with – the original Fender-sound allegedly can only be achieved with the ABC-Orange-Drops, while the opposing camp warns of using exactly these ABC-Orange-Drops (mid-rangy sound, totally unsuitable), recommending rather the yellow Mustard-caps? Wait – not the ABC-Mustards: these are inadequate copies (clones, so to say), you should use the others, the original copies. Even better: use silver-foil capacitors, or, if that much budget can be committed, copper-foil-caps. … "committed" …, no let's not go there, and rather focus, with the naïve curiosity of the researcher/scientist, on the task at hand: trying to find for a grain of truth in that pile of rubbish.

In the framework of the present considerations, a capacitor is a component in an electrical circuit, and as such is subject to the rules and standards of electrical engineering. Whether it has an aura, whether it holds spiritual energy or ethereal psi – that will not be investigated here. The very powerful instrument of **Maxwell's equations** describes, to general satisfaction, the processes in electromagnetic fields. It has made wireless transmission, space exploration and EMP computable; so why not use it as well on the triode-preamp-stage of a guitar amplifier? These equations are the big guns pressed into action whenever starting-from-scratch is called for – but fear not, they will serve here as a mere launch pad that is left behind as quickly as possible. With the limitation to the audible frequency range and to concentrated components (i.e. components smaller than about 1 km in size), Maxwell's second equation may be simplified, resulting in Kirchhoff's second law (the **loop rule**):

$$\oint_K \vec{E} \cdot d\vec{k} = 0 \quad \Rightarrow \quad \sum_{i=1}^{n} U_i = 0$$

Maxwell's and Kirchhoff's 2nd law, resp.

The line integral of the field strength $\vec{E}$ along the closed curve K is zero, as is the sum of the n branch-voltages U_i along the closed loop. Correspondingly, the (slightly modified) 1st equation of Maxwell is:

$$\oiint_A \vec{S} \cdot d\vec{a} = 0 \quad \Rightarrow \quad \sum_{k=1}^{n} I_i = 0$$

Maxwell's and Kirchhoff's 1st law, resp.

The enveloping-surface-integral over the current density $\vec{S}$ is zero, as is the sum of all n node currents I_i (**nodal rule**, rule of charge conservation).

In an electromagnetic field, there are just three quantities characterizing the material: the specific resistance ρ, the permittivity (dielectricity) ε, and the permeability μ. Limiting ourselves to the audible frequency range, we may disregard the magnetic properties of typical materials used for capacitors. Thus, for the formal description merely the loop rule, the nodal rule, and two equations relating to the material remain. Despite this simple analysis, a number of misunderstandings exist that have their roots in the inappropriate application of actually appropriate insights. The area of electrical engineering was in fact not developed to build guitar amplifier but it is, with all due respect, quite a bit older. Besides other important technical fields such as power engineering, one main focus was communication technology, and here in particular the question how to achieve long-distance **wireless** transmission of speech or Morse-code. Heinrich Hertz [1886] and Guglielmo Marconi [1901] are to be mentioned, amongst many other pioneers. The first radio transmission across the Atlantic succeeded a generation before the start of the Rickenbacker/Gibson/Fender-age, and already then capacitors were in use. The so-called "Leyden jar" was invented even much earlier in 1745. Radio transmission, however, does expressly not work in the *audio* frequency range. High frequencies (or *radio* frequencies – sic) are required: 1 MHz for the AM-range and about 10 MHz for the short-wave range. Considering this, and also the fact that the "bibles" from back in the day were titled "The Radio Engineers' Handbook", and not "The Guitar Amp Designers' Handbook", it is easy to imagine what can happen: someone (from the guitar community) reads that caps may have their problems at higher frequencies, and immediately fears for the silvery highs of the famous Fender-sound. As if they had been built for just that situation: there are indeed **Silver-Mica-Caps**! They will make that Fender-treble bounce right back, won't they! The echoism "silvery highs" should not be criticized – that can very well remain here as a term of art. It is also correct that capacitors become **inductive**, but at which frequency does that occur? Even for a wound capacitor (very remotely related to a coil), effects of this inductance happen only above 1 MHz, i.e. at "higher frequencies". However, that does not mean the "higher *audio* frequencies" are affected – there is a factor of about 50 separating these ranges! In just the same way, the **loss-factors** known to "rise towards higher frequencies" will worry only the RF-technician or HF-filter-designer. Typical guitar amplifiers, however, include neither a short-wave-input nor a 12th-order elliptic filter, and the low-loss styroflex capacitor is rather misplaced here (it's too hot an environment for it!).

On top of the difficulties to interpret books on RF-technology in the correct way, we also encounter the problems found in **psychometrics**: how do we measure audio-perceptions? In Chapter 10.9.4 (sound event, auditory event), some suggestions were made, and in particular the need for blind-tests was emphasized. Daily practice, however, looks quite different. For example, a manufacturer states that his capacitors require a run-in time of **100 hours** until they sound good. The guitarist having fitted his (sound-wise not convincing) Marshall with these caps hears exactly that, and reports to all colleagues: "it was only after 100 h that the treble was how it should be." Indeed, it is generally known that technical devices require a run-in time: that relates to the car-engine (1000 km run-in) as it does to the charcoal on your grill (no smoke should remain after 10 minutes. And if a newly elected politician gets a 100-day grace period: why not the capacitor, as well? Let's consider this: an amateur musician playing for 10 h a week will after 2,5 months arrive at the point where he remembers how the caps sounded when they were just freshly soldered-in: it was atrocious, the treble just wasn't there. Now, however, after almost a quarter of a year, it suddenly has appeared. And the reason for this improvement must not be attributed to the tube-aging♣, not to the (possibly new) loudspeaker and certainly not to the strings changed multiple times. No, it must be the caps – in fact the guitar mag has found the same, and even recommends 200 h of run-in time

♣ the manufacturer says (1952 A.D.) that a run-in time is required. Why are we not surprised: it's 100 h.

(i.e. … wait another quarter of a year). And why wouldn't it be the caps – for tubes, nobody doubts time-variant sound changes, either, do they? Whether copper-foil capacitors sound better than aluminum-foil capacitors ('cause copper has lower resistance) – somebody must be able to verify that beyond reasonably doubt! Skeptics pointing to missing data regarding the thickness of the foil could possibly be convinced, if indeed the listening experiments would not show such dramatic deficits. If such an experiment includes the announcement: " I will now solder-in the copper-foil-caps; that will give a much more forceful sound", it immediately becomes futile since all participants will be biased. A much better approach can be found with **Tone-Lizard.com**: here it may happen that the same capacitor is connected to 5 taps of a 6-position switch – not 5 equal caps but in fact one and the same capacitor connected to the 5 positions, and without the judging guitarist knowing this. The latter declares: '*the orange-drop in position 1 sounds much better than the one in position 4*'. No further comment necessary. Obviously, this is just a single case, but just as obviously, the judgment ' *I cannot for the love of it hear any difference*' is not likely to occur a lot because no sound-expert will want to damage his/her reputation. And so capacitors all have their own sound, as does every bolt and every rivet.

It is not possible to find out who first introduced the myth of the 100-h-run-in-time for capacitors. The following might be an explanation, though. In data sheets for capacitors we find a **time-constant** of 100 000 s for high-grade builds – this equals just under 28 h. If now someone has some lingering remains of memory that for a full charging process, 5 time-constants waiting time is a good measure, then we arrive at 140 h … there it is – the run-in time? No, that's not it, at all, of course – these are all data relating to the insulation (!!) behavior (which is not unimportant but an entirely different issue, so let's postpone the discussion of it a bit). The above <u>idle-state</u> time-constant would only play any role for the capacitor by itself i.e. completely disconnected (which you wouldn't want in a guitar amp, would you!?). Connected in its regular habitat, the capacitor is loaded on one end with the grid resistor (e.g. 1 MΩ), and at its other end with the tube and the plate-resistor (e.g. 50 kΩ). Multiplying these resistances with the capacity yields the actual time constant in operation, and that is, at e.g. 22nF · 1,05 MΩ = 0,023 s, significantly smaller than the idle-state time-constant. If you really could talk about a "run-in time" here, and multiplied (according to the above approach) that time by a factor of 5, you would arrive at a "run-in time" of a full 0,12 s.

Alternatively, there is reasoning based on material changes that start only after the onset of a voltage and need to stabilize first. It appears that someone has looked too long over the shoulder of the colleagues dealing with **electrolytic capacitors**. This type of cap uses, similar to the foil-capacitor, a rolled-up aluminum foil – however the dielectric is not formed by plastic foil between the aluminum layers but by a thin Al_2O_3-layer (aluminum-oxide) that grows onto the anode of the capacitor during a **formation process**. It is only this oxide-layer that has the insulating effect, and therefore an unformed electrolytic cap must never be subjected to a voltage. However: what does this have to do with the coupling capacitors that are never, ever, of the electrolytic sort in a tube guitar amp? In the latter, polyester or polypropylene capacitors are used that do not have to – and cannot – be 'formed' in the first place. Still, couldn't there be any other slow-moving and possibly unknown processes at work on the metal surface or in the dielectric? Yes, actually, that is highly likely – the field-strength is high and the temperature, too. Let's assume that a capacitor somehow changes during the first 100 d. What parameters could in fact change? In any case the **impedance**, i.e. the complex ac-resistance, would have to change. Only variations that would have any effect at all on the impedance could change the grid-bias-voltage – hopefully nobody will claim that the sound might change while all tube voltages remain the same. Now, here's the thing about impedance: it is only defined in the linear model; there are no impedances in the **non-linear**

model. Could a coupling capacitor be non-linear? Sure, every capacitor is non-linear – it increases its capacity when a voltage is applied (charged electrodes attract each other, decreasing the distance between them and increasing the capacitance). However, already simple math shows that the resulting pressure (some kPa) remains (together with the elasticity module of around 1.000.000 kPa) below the critical deformation limits by several orders of magnitude. Measuring the THD confirms: as long as a HDK-capacitor (highly uncommon in guitar amps) is avoided, the THD remains below 0,01 %. Thus: we do have linearity in caps – with good approximation.

So: which parameters could change? To start with the **insulation resistance**: it could e.g. increase from 50 GΩ to 100 GΩ (☺), or it could decrease to e.g. 25 GΩ (☹?). Taking 1 MΩ as the input impedance of the following stage, a leakage current of 4 nA (or 2 nA, or 8 nA) would flow at an operating voltage of 200 V, yielding a shift in the operating point of 4 mV (or 2 mV or 8 mV). None of these values represents any reason for concern, plus (most importantly): at an insulation resistance of 25 GΩ, the manufacturer would have pulled the plug because this would be outside of the specification at least for branded products. Incidentally: will you wait for 100 h for the insulation resistance to deteriorate? That could be achieved in a much simpler way. In any case, the insulation resistance may be a shambles (as it actually is for the very long serving caps in vintage amps – sic!), but for new name products it will be good enough, and will not have any audible effects on the sound.

And on to the **capacitance**. Does that change during the first 100 h, right? Of course it does - panta rhei – everything changes all the time. Data books for high-quality capacitors indeed specify a change: e.g. <± 1%, at 70°C , for the first two years. No cigar, then, either. And again we need to note: if the sound were indeed better only after the capacitance has in- or decreased by x% - the improvement could much simpler be achieved than by waiting for all that time.

The **loss factor** remains. It could drop – that sounds desirable even without any exact knowledge (lower loss of whatever) – or it could rise. Hold on … the amp sounds better after 100 h, so presumably it will be *lower* losses? From "smeared sound to clear sound", as the guru elucidates. Has anyone actually posed the question why we would solder a capacitor – no: several capacitors for serious money into our amp, if it sounds “smeared” for 100 or 200 or even more hours? Someone probably did ask, and the stunning answer would be: “because afterwards a undreamt-of sound experience will set in that off-the-shelf products can never impart.” Per Aspera ad Astra – we know this, as well. If the loss-factor were crucial to the sound (and indeed who wants any “loss” in their sound), then the lowest-loss dielectric would have to be used in coupling capacitors, correct? Consequently, and after excluding the les temperature-resistant polystyrene, **polypropylene** would be the right choice. And indeed, it is exactly this material that is found in the Orange-Drop-caps already mentioned. However: the guru from the other side of the fence strongly advises not to include those, but only the ones with **polyester** as dielectric, just as in the original in the 1960’s. The manufacturer adds: “because of its deeper tonal quality" – whatever that may be. Polyester, however, has a loss-factor about 100 times higher than of that of polypropylene – so now the going gets tough. Could the solution be: *more* losses for a better sound? A short calculation comes to the rescue: polyester will give a loss-factor of a about 1% at 10 kHz which – in the high-frequency equivalent circuit (to be discussed later) corresponds to a 7-Ω-series-resistor (for C = 22 nF). Imagine this connected in series with the source impedance of the preceding tube stage (e.g. 50000 Ω): it will be 50007 Ω instead of 50000 Ω – so much then for that approach. Clear? Or smeared?

No, the losses do not help us here. In short-wave resonant circuits, there would possibly have been trouble, but nobody will include a 22-nF-capacitor is such a design. Nope, losses do not explain any sound difference in coupling-C's.

Don't stop now – here comes the **slew-rate**: the cap-manufacturer still has some aces up his sleeves. The slew-rate is the differential quotient over time of the voltage across the capacitor; it is also called the speed of voltage-change, given in V/µs. Again, a glance into the data books is helpful and indeed they show limit-values that must not be exceeded. Examples are 500 V/µs for a polyester capacitor, or a mere 30 V/µs; or 750 V/µs for the polypropylene cap, or even more than 1000 V/µs. Not to forget: the *mica capacitors that considerably improve the sound* excelling at 100.000 V/µs. These are considerable differences, so what is the secret? Let's do the math: a tube that is required to generate an AC-voltage of $30V_{eff}$ at the plate reaches 0,8 V/µs. (N.B.: we work with 3 kHz here since the pickup will never manage full drive-levels at 20 kHz). Or maybe a bit more which brings the slew-rate to just about 1 V/µs. Given the above data, this would not cause any problems. However, the tube may be dramatically overdriven, e.g. by a factor of 30 – and now we would arrive in the range of the above limit-values. Strike!!?? No, still not: the slew-rate happens at the plate but not across the capacitor! The voltage across the capacitor is in fact much smaller, and it even decreases with increasing frequency. That's why the slew-rate is not a meaningful parameter – the **maximum current load** that can equivalently defined for the cap would be more suitable. The data books specify values of 0,1 – 1 A, and even *more than 1000 A for the mica capacitor* – is that good enough? Well, yes – your typical noval-triode is able to supply only a few mA. So again: no issue. But here's a hunch based on this line of thought: the originator of the buzz around the cap might have built a loudspeaker crossover at some point. That scenario provides an entirely different picture – we are confronted with the big-boy-currents: $100\,W = (5A)^2 \cdot 4\Omega$, i.e. 7 A peak current. Could it be that someone has put to (mis-) use the **frequency-crossover** design-rules in the context of coupling capacitors? Sure: the manufacturer of the expensive copper-foil-caps is also known for his crossovers, isn't he? Directly quoting the guru: "it's an unbelievably fat sound …" Unbelievable?

An on we go: the **inductance** of a coupling capacitor now surfaces. It may *today be reduced to the point where 0,8 nH/mm CS can be reached.* CS stands for "contact spacing" and not for "compact size" … but the latter would anyway not be appropriate given the sheer size of these super-coupling-caps (45 mm x 26 mm). In the Roederstein (ERO) data sheets we find much larger Nanohenry-values (12 nH) – but STOP: that brochure dates back to 1980. What was the size of that super-cap again? 45 mm in length? So it's 45x0,8 nH = 36 nH? And ERO was also per CS? No, that was the overall inductance. My, so much has changed since the 80's! Just to mention it: even 36 nH would be o.k. – at 10 kHz a reactance of a full 0,0023 Ω would result. That would have to be added to the 50000 Ω mentioned before. Pythagorean addition to be used, of course.

So, what remains? Not much … maybe the skin effect: that is also an object of adverts. "*At high frequencies* (ah, the old story …) *the current flows merely along the outside of the wire, such that the conducting cross-section is reduced and the resistance grows.*" That is entirely correct: *at high frequencies*. H. H. Meinke always started his famous lectures with just this issue, and noted: "… *and that can start already at 10 kHz*". It is from around that frequency that the resistance of connecting wires increases noticeably – maybe from 0.002 Ω to 0,004 Ω. Nothing could be added here that was not already covered above when we looked at the loss factor. No, the skin effect doesn't contribute anything either, at audio frequencies. And another thing: someone advertises that the connecting wires of his caps are made of tin-coated copper, and not of copper-coated steel reaching merely 30% of the conductivity of copper.

Here's a suggestion for improvement: "*our replica caps come including the connecting wires and not without – the corresponding enhancement in terms of conductivity is beyond anything that could be expressed as a percentage.*"

Lest our commentary turns into mere satire, let's try to arrive at a summary. First, however, a look into the internet: word is that there are rationales that are so misguided that it is not possible to even get close to them on the basis of normal training in electrical engineering. Our capacitor-guru provides some inspiration: *every capacitor brings* ***unwanted effects*** *in the form of additional inductances and resistances (ESR = equivalent series resistance). While the inductance practically has next to no influence, the ESR determines – in conjunction with frequency and capacitance - the loss-angle tan(∂).* O.k. – it is possible to express the issue that way – all text-books on electrical engineering do it in a similarly. Let's move on: *tan(δ) does not remain the same for all frequencies but is frequency-dependent* (that is correct). *If tan(δ) rises strongly with frequency, the frequencies are not all treated in the same way; if there is a smaller change between 1 kHz and 100 kHz at least the frequency spectrum of an impulse mixture is transmitted more time-correctly. The highest frequencies are shifted in terms of amplitude and phase = differential phase-error. For the ideal capacitor, tan(δ) needs to be as frequency-independent as possible implying an ESR that drops towards high frequency. That way, time-distorted frequencies are shifted further towards the MHz-Range (=smallest possible error within the audio-range).* Phew – now we have arrived at a hodge-podge of desire & reality, of science & sales – this is now really incorrect.

The above "sales-supporting comment" could be shortened to: *every signal-carrying capacitor deteriorates the reproduction of impulses, and it does this the stronger, the more the loss-factor depends on the frequency.* Whether there are actually any time-distorted frequencies does not need to be investigated yet, but the reproduction of impulses needs to be looked into. Let's start with an equivalent circuit for a simple high-pass, as it is also presented by our capacitor-guru for a coupling capacitor (**Fig. 10.9.7**):

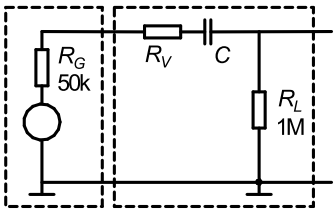


Fig. 10.9.7: The coupling capacitor in the high-pass circuit.

The source (this could be the plate-circuit of the preceding tube) is drawn as a voltage source with a series resistance R_G, R_V = ESR, the load (this could be the grid circuit of the subsequent tube) is shown as a simple real resistor R_L. If needed, this simple schematic could be extended, but for basic considerations, it is adequate. R_V is frequency-dependent – our guru states this, as well. He does not state that in this equivalent circuit diagram, C needs to be frequency-dependent, too – however, since this dependency is small, it may be ignored for now (just as L may be ignored for now). We assume that R_V is a simple, real, frequency-independent resistor, for example a short length of copper wire. Copper? No, why don't we better turn to silver wire as it is offered likewise by the capacitor manufacturer: with a choice of Sterling silver (92,5%), or pure silver (99,97%). For our thought experiment, we naturally take only the purest silver because of the higher conductivity (silver conducts about 5 – 10% better than ordinary copper). Such a short piece of wire will measure maybe 2 mΩ, and this we have to think as connected in series with our source impedance. The two resistors maybe added up yielding a full 50.000,002 Ω. This is one of the undesired side-effects: the source resistance is increased.

What does the loss-factor of this model-capacitor look like? Since the ESR was assumed to be frequency-independent, tan(δ) rises proportionally with frequency. Big mistake, because: *for the ideal capacitor tan(δ) needs to be as frequency-independent as possible,* i.e. the ESR should decrease with $1/f$: $R_V \sim 1/f$. The guru states: *the smaller the increase of tan(δ), the higher the voltage steepness dV/dt and the smaller the differential phase-errors in the audio range – especially for higher* voltages. Voltage-dependent errors? The model of our guru is a linear one and thus cannot include any voltage dependent components. It isn't that capacitors would show no non-linear effects at all, but if you want to describe non-linearity, you have to have a non-linear model. Let's stick to the linear model, though, in order not to further increase the already considerable confusion; for small voltages things are linear, anyway.

Let's go back: we have the allegation that *highest frequencies are amplitude- and time-shifted.* Just to be precise: a frequency cannot be shifted – neither in amplitude nor in time. Frequency is the inverse of the cycle duration: a cycle duration of 10 ms results in a frequency of 100 Hz. What is presumably meant: highest-frequency signals are time-shifted and changed in their amplitude. We could also say: in the highest frequency range, amplitude- and phase-changes occur. That, indeed, is how this would be expressed in systems theory. *Differential phase-errors* – no, you wouldn't really say that. What could be meant here? Maybe it is the spectral derivative, $d\varphi / df$, that usually is supplemented with -2π and designated the **group delay** $\tau_g = -d\varphi / d(2\pi f)$. This is in systems theory, usually. It is not problematic to use uncommon terms, but for misinterpretations, the liability needs to go to the party responsible. Again, in a nutshell: *the ideal capacitor does not generate any amplitude- or delay-distortion*♣. Is that what was meant? O.k. then – let's move on.

Not any amplitude-distortion. At another passage it is even more drastic: *the ideal capacitor should not influence the audio signal at all.* So obviously, there must be signals that are not audio signals, and these are clobbered by the capacitor? Correct? No – wrong! In a loudspeaker-crossover, the capacitor connected ahead of the tweeter is supposed to attenuate low- and mid-frequency signals i.e. it is there to reduce their amplitude. Are these signal also audio signals? Of course they are! But lets set aside the crossover – it is indeed possible to talk, in the context of a coupling capacitor, of *small influences on the audio signal, or none at all*. The capacitor operates (together with the resistors) as a high-pass filter and therefore attenuates the (very-) low-frequency components. Our guru, however, seems not interested at all in the low-frequency range, since he localizes – even several times – the undesired effects in the range above 1 kHz: *a capacitor would let high frequencies pass without limitation if it weren't for the losses.* Herein lies a grain of truth: the impedance of a capacitor would continuously drop with increasing frequency if there would be no losses. But wait a moment – is this really about the **impedance**? At 1 kHz, the losses (10^{-4}) increase the impedance of a 22-nF-cap (polypropylene) by 0,0000005%. That's no joke, it's covered in the basic course in material science and components => $\sqrt{1+\tan^2(\delta)}$! Even at 10 kHz (and further up in the audio range), the impedance is not an issue. So, then, on to the **phase**, or rather its spectral derivative. We read that it would be ideal if *voltage and current would occur at the capacitor with an exact 90°-phase-shift between them*. Indeed, that is correct: only the ideal capacitor can achieve that.

♣ In systems theory, the term *distortion* is employed in two ways: non-linear distortion (harmonic distortion) and linear distortion (amplitude-, phase- and delay-distortion).

The real, lossy capacitor shows a phase-shift that is different from 90°. Is that bad? Apparently so, because *differential phase errors in the audio range* result. If the phase shift is frequency-independent, then "all frequencies arrive at the same time" – to stick with the terminology of our capacitor-guru. In other words: signals of different frequencies all are subject to the same delay. However, if the phase-shift is frequency-dependent, the "high frequencies are time-shifted". Or better: higher-frequency signals are time-shifted (or delay-shifted, or, alas, phase-shifted) relative to lower-frequency signals. Possibly, an impulse will be stretched out, or "smeared", due to the phase-shift. We do recall that term: "smeared sound". The absolute phase difference between current and voltage (which for the ideal capacitor will be 90°) can be accepted; the *differential phase-error* (that, in the absence of any explanations, needs to be interpreted as the group-delay) causes impulse distortion. So, our guru had a convoluted and sibylline way of saying it – but what he actually wanted to express is this: **"The loss-factor of a capacitor causes group-delay distortions that smear the sound (at high frequencies)"**. As a remedy – direct quote guru – the loss factor needs to be as independent of frequency as possible, or, in other words, the group-delay needs to be as frequency-independent as possible. At first glance, this sounds familiar: systems theory says such a system is a **linear-phase system**, and certifies a distortion-free behavior.

And with this, we have arrived at the core of this grandiose misunderstanding: the group delay is a transmission quantity (a quadripole-quantity), while the phase-shift between current and voltage is a two-pole quantity. Differentiating the wrong phase will yield a wrong result. More specific: a **quadripole** is a system with four terminals (also called two-port network), with a two-terminal input and a two-terminal output – the high-pass shown in Fig. 10.9.7 would be an example. That input and output in this example have the same ground connection does not make the system a tri-pole – it still is designated a quadripole. Between the input signal (input voltage) and the output signal (output voltage), a complex transmission function is defined from which the frequency response of the phase and of the group delay can be derived. A capacitor, on the other hand, is a **two-pole** because it has merely two terminals. A complex impedance is defined between the voltage and the current, and a phase frequency-response can be derived from this impedance. But try and deduce a group-delay frequency-response from this – that is nonsense. There are merely two special scenarios in which it is purposeful to see a two-pole as a quadripole: if voltage is the input quantity and current is the output quantity, or vice versa. Now, one could argue that every quadripole is on fact constructed from two-poles and that therefore any deficiencies of these two-poles must also be a deficiency of the quadripole. This, however, is not the case. In the present framework, we cannot present the systems theory in its full scope but have to refer the reader to special literature [5, 6, 7]. Very briefly: in the high-pass mentioned above (and equally for an RC-low-pass), the input voltage is divided between R and C – it does not fully span across the capacitor.

The simple formula $d = \tan(\delta) = 2\pi f \cdot R_V \cdot C$ yields the loss-factor d as tangent of the loss-angle ∂. In the ideal capacitor, a sine-shaped voltage precedes the current by exactly 90°; for the real capacitor this angle is smaller than 90°. For a loss-angle of $\delta = 0{,}01°$ (polypropylene at 1 kHz) the phase-shift between current and voltage therefore does not amount to 90° but to 89,99°. If the loss-angle were frequency-independent, as curiously demanded by the statements of our guru, then the phase-shift differentiated with respect to the frequency would result in a constant value of zero (the derivative of a constant is indeed zero). That would seem to be the ideal case: no smearing of impulses.

Now, the phase-shift between the current through and the voltage across a capacitor is one thing, and the phase-shift between the input and output of the coupling circuit is something different – in fact something entirely different.

The phase-shift φ appearing in Fig. 10.9.7 between the generator voltage and the output voltage (across R_L) can easily be calculated:

$$\varphi = \arctan\left(\frac{1}{(R_G + R_V + R_L)\cdot 2\pi f \cdot C}\right) \approx \frac{1}{(R_G + R_V + R_L)\cdot 2\pi f \cdot C}$$

[6, 7, 17, 18, 20].

Fig. 10.9.8 shows the frequency response of the phase in a presentation with both axes in logarithmic scaling – for the ideal, loss-less capacitor! In reality, the group delay♣ depends on the frequency, and dispersive impulse-distortion results. Again: the case calculated here is the best-possible one with the phase-shift between voltage across and current through the capacitor being exactly 90° at all frequencies. However, for a real polyester-capacitor, practically the same figures would emerge – the difference would be entirely insignificant: e.g. for the group delay it would be as little as 0,0004% (1 kHz). Using a capacitor with a constant loss-angle across the frequency range would deliver differences of a similar magnitude. Relative to a load resistor of 1000000 Ω it does indeed not make any difference whether the ESR is 7 Ω or 0 Ω. What does make a difference is a change in the capacitance – shown in the figure for a 30%-increase. It will be discussed later whether such huge **tolerances** can occur at all in a high-grade capacitor, and, if yes, whether they are significant.

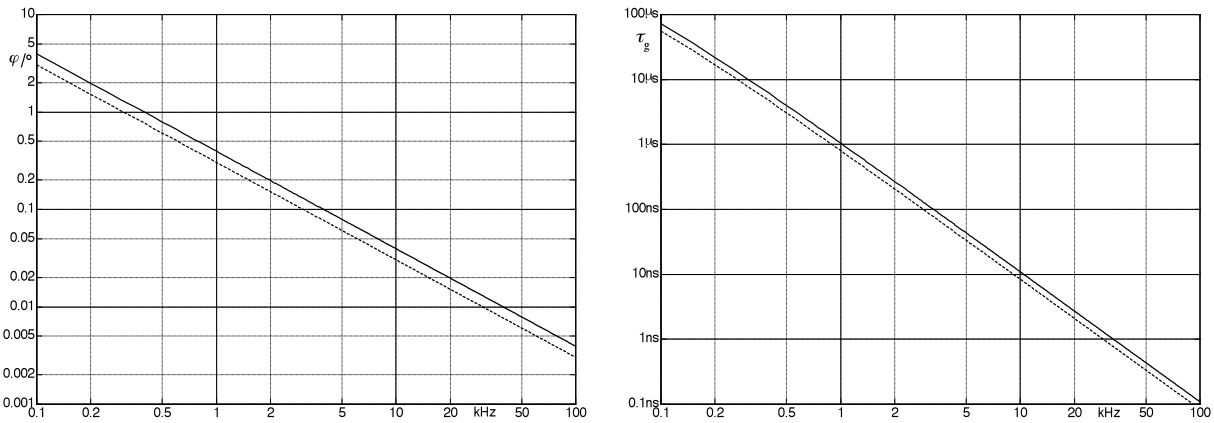


Fig. 10.9.8: Phase frequency-response and group-delay frequency-response of the circuit acc. to Fig. 10.9.7 R_G = 50kΩ, R_V = 0, R_L = 1MΩ, C = 22 nF. The dashed line is valid C = 28,6 nF (i.e. +30% tolerance).

In summary: the conjecture that the loss-angle would have to be as frequency-independent as possible leads to incorrect conclusions, since it is derived from an entirely unsuitable two-pole phase-angle. For a typical coupling circuit with tubes, all capacitors (including theoretical, ideal capacitors) generate practically the same group-delay distortion ("impulse smearing"). This delay distortion is, however, so small, that it remains far below the threshold of audibility.

♣ In case the group-delay is to be derived from the phase frequency response via graphical differentiation (gradient), a representation with linear scale on both axes needs to be used, not with double logarithmic scale.

The last statement needs to be commented on. First, we give the word to our guru: *the value of tangent ∂ does not always tell the full story. If it rises drastically, not all frequencies are treated in the same way; if it changes less between 1 kHz and 100 kHz (as it is usually the case for metal-foil capacitors), the frequency spectrum of an impulse-mixture should be reproduced more accurately with respect to time. In this context, a shallow rise of tangent-∂ between 1 kHz and 100 kHz is desirable, combined with a simultaneously stronger dropping ESR. For the same reasons, developers aim for rise-times of below 1 µs (about 1 MHz bandwidth).* That is plain wrong. The term "rise-time" is used in circuit technology to define the time period during which an impulse response rises from 10% to 90%. The rise-time is 2,2 times the length of the time-constant that in turn is the inverse of the angular cutoff-frequency. From this, the cutoff-frequency calculates as $2{,}2 / (2\pi \cdot 1\mu s) = 350$ kHz, i.e. not 1 MHz. Even if not rise-time but settling-time were meant, the bandwidth-specification is still incorrect: for 1 µs settling-time, a bandwidth of 500 kHz results for steep-slope bandwidth limiting (since the term is not really used much for a first-order low-pass). Anyway, rise-times of below 1 µs, since: *the superposition of room-reverb onto the original signal needs to be correct to the microsecond, so the ear can pick out the exact location in the space.*

This we need to think about – that is not entirely wrong. The threshold for detecting an inter-aural delay (localization blurring) can be determined to be as low as 10 µs under laboratory conditions, and specialist literature reports even lower values than that. And if you really want to achieve a "safety zone" of a factor of 10, the result is indeed 1 µs. However, to conclude from this the requirement of a bandwidth of 350, or 500, or even 1000 kHz – that would be nonsense. A pure 1-kHz-Tone can perfectly be shifted by 1 µs without tapping into the RF-range. The hearing system can perform the delay-resolution of 10 µs (as mentioned above and if indeed it does that well at all) in the mid-frequency-range, i.e. at around 1 or 2 kHz. At 10 kHz, this just noticeable difference has grown quite a bit (100 µs as a rough guideline – the data depend highly on the experimental conditions), and beyond 20 kHz there is no hearing. Or is there?

Now, every audiophile has gathered (from wherever) that the stimulation with pure sine-signals is something quite different than real sounds because the latter contain tons of impulses. And so one of our gurus manages to demand, on his webpage, on the one hand a bandwidth of 1 MHz, and to refer, on the other hand, to a thesis that very accurately takes the upper limit of hearing to be at around 19 kHz. How does that fit together? We are not talking about 19 or 20 or, even better, 22 kHz – here very casually a factor of 50 is built in, as a reserve. To voice, in one and the same sentence, an opinion and simultaneously the counter-opinion – that is normally only achieved by certain politicians (or showbiz-people).

This mixing-of-what-must-not-be-mixed-up is done – for audio signals – in the following way: *every signal – and that means indeed EVERY signal – is in fact the sum of an infinite number of sine signals*. Yessss!! You can't maneuver around good ol' Baron Fourier. In principle, this statement is correct but we must not take the "in fact" too literally. Mind you: the Fourier analysis is a model consideration, and every signal could just as well be segmented into many other (not even necessarily orthogonal) functions rather than sine functions. What is valid for signals is also valid for systems (as long as they are liner and time-invariant): the consideration of processes in the spectral domain is equivalent to the consideration of the processes in the time-domain [6, 7, 17, 18, 20]. If the hearing system cannot hear continuous tones with a frequency of above about 20 kHz (and moderate SPL), it cannot hear, for impulses, any of their spectral components that lie above about 20 kHz, either.

Now, this finding must intricately be reformulated in such a way that the capacitor manufacturer will get a sales boost. That is done in the following way (under the same web-address): *music is not just composed from pure sine-tones but from a very broad spectrum of different impulses that in part lie very far outside the hearing range but which strongly influence the hearing perception*[⊗]*. For this reason we use – for audio amplifiers – a capacitor that falters only above 100 MHz. In the coupling branch of a pre-amplifier, this capacitor gives us an impression of how a piece of wire would sound.* No, this is not a printing error – it indeed is supposed to read not 100 kHz but 100 MHz. And a few lines further we find: *mica is the ideal dielectric for capacitors, yielding the following properties: ... applicable up to very high frequencies into the GHz-range. Mica capacitors are highly favored for filter circuits where, due to their properties, they can bring a considerable increase in sound-quality.* Gigahertz! Is there no end to this?! Is the Terahertz range next? And yet Schöne et al. have already proven in 1979, that a reproduction of the ultrasound range adds nothing whatsoever to the perception[♣]. That was an investigation carried out by the Institut für Rundfunktechnik (the internationally renowned German broadcast technology institute), though, and in some audiophile circles the preference is not to take note of research done there. Any self-appointed guru who pushes the requirement a further few MHz into the RF range is seen as the new messiah. Skeptics, however, are branded as "*infidel physicists whom one should give a wide berth*". "C'est la gare" is the only congenial answer to that.

Let us revisit the example used in Chapter 8: a bed of a length of 1,5 m will be judged as too short for most grown-ups, while a length of 2 m is quite comfortable. Now, there are a few people who are taller than 2 m, and to accommodate these cubo-philes, a bed should, for good measure, be a bit longer. Taking the above approach used by our capacitor manufacturer for the 100-MHz-capacitor, the bed should be about 10 km long, just to stay on the safe side. Has anybody thought of "indulging" our other senses that way? Our visual sense would lend itself as a candidate: the limitation to the frequency range generally as "visible" (380 – 770 THz) seems overly restrictive, and why not give the TV a correspondingly enlarged bandwidth (i.e. X-ray radiation)? And, of course, that should extend into the lower range, as well: the microwave oven would stand ready to be a splendid "optical subwoofer".

But back to the audio amplifier: the frequency range up to 20 kHz needs to be reproduced precisely, and since no amplifier will shut down abruptly above this limit, a few more tens of kHz are purposeful to let the amp taper off. In case listening experiments result in other numbers, any conclusions may be put under scrutiny on the test bench. Ill-considered phase responses and listening experiments with biased subjects are, however, not conducive. And one more thing about the 10-µs-delay-distortion mentioned above: even smaller values may be audible inter-aurally. Given that, and the fact that most people have evolved beyond mono into stereo-territory, wouldn't it be desirable that the capacitance-tolerance of the wonder caps would be of matching dwarfishness? From this point of view, it is peculiar that one of the manufacturers specifies tolerances of -20/+30%. Sure, hand-made, every capacitor is one of a kind. Or maybe the manufacturer is aware how strongly the group delays of any two headphone systems or of two loudspeakers (of the same type, respectively) can differ? Maybe he knows all this and just doesn't tell? And continues to jumble and confuse things while feverishly searching for the ideal capacitor that blocks and at the same time passes DC.

[⊗] That is why they are called "lying outside of the hearing range" (sic).

[♣] P. Schöne et al.: Genügt eine Bandbreite von 15kHz... (Rundfunktechnische Mitteilungen, 1/1979).

From the world of the space-worthy 100-MHz-capacitors that can deal with 1000 A now back to the guitar amp. Don't panic: here we don't find either. Even the 10 µs that may be crucial of the inter-aural delay do not give us a headache. For all guitar players that do not carry a stereo-system to the stage: the threshold for diotically♣ perceivable group delays is about 2 ms [3, 12]; one may get to somewhat smaller values via special training, but this is not an issue in practice. So, for sure, there is no "smeared sound" due to the coupling cap with its group delay of τ_g = 0,0001 ms. Things may be entirely different in loudspeaker crossovers where we have large currents on the way – maybe not 1000 A and 100 MHz but still: this is the power-engineering-league. The coupling capacitor plays in Little League: it's communication engineering and a few microamperes on this playing field.

Let's acknowledge the difference (whatever it may be) between power- and communication-engineering, and between research and marketing. After we have (full-monty?-) scientifically shown that coupling capacitors cannot contribute anything – really not anything – at all to the sound, we could conclude with a real bombshell and note that these caps do in practice influence the treble, after all. In fact, that is easily explained and we will get to its in a bit. First, the relation to the equivalent circuit needs to be covered, in more detail. Gotta do it.

In the daily routine in the lab, a coupling capacitor is described via two quantities: capacitance (e.g. 22 nF) and dielectric strength (i.e. proof voltage, e.g. 400 V). The third parameter (the loss factor, is of significance only if the capacitor is connected to inductances. This would be the official position, and according to it all capacitors of equal capacitance would have to sound the same. The teachings of electrical engineering do however also state that the function of a capacitor is of such infinite complexity that only rigorous simplification makes the above analytical description possible. The series connection of an ideal capacitor and an ideal resistor is just about the simplest approximation: more sophisticated models consider special polarization effects, as well, and they arrive at more complex equivalent circuits (Chapter 9.4). However, at middle and high frequencies coupling capacitors are of such low impedance that only a very small AC-voltage is created across them. Of the 30 V_{eff} plate-voltage, a mere 0,2 V_{eff} are found across a 22nF-high-pass capacitor (22nF/1MΩ) at 1 kHz, and therefore the divergence of that cap from the ideal cap is not that significant. At low frequencies, however, the AC-voltage across the capacitor rises: half the frequency – double the cap-ac-voltage, until the cutoff frequency is reached at 7 Hz. Somehow, though, the lows never turn up in reports about the sound of coupling caps; it's always the highs that are smeared, that sound "mushy" or "hollow", and only "open up" after 100 h. Still, we could – for once – consider the lows as well … the real deep lows:

Let us consider once more the high **DC-voltage** across coupling capacitor: depending on the circuit this will be 150 – 300 V, in special cases even more (beware: mortal danger!). If the insulation resistance of a coupling capacitor is e.g. 1 GΩ, about 200 mV are measured across the following 1-MΩ-resistor (at 200 V plate voltage) – for an ECC83, this is already quite a lot (Fig. 10.1.14) and may cause audible effects. However, whether the sound is improved or damaged by this offset-shift cannot be generally predicted. There is always the same rationale: we encounter too many sound-determining parameters. This lack of a general prediction may not be really necessary, any way: for new high-grade capacitors, the insulation resistance is far higher than the one used in the above example, but for decade-old capacitors it many be much lower.

♣ diotic presentation: both ears receive the same signal (mono, both ears listening)

In data sheets we find, for polypropylene-caps, insulation resistances specified to 20000 GΩ, and of 1000 GΩ for MKT-capacitors (each for 22 nF). These values are given for room temperature – which is not the normal internal condition in a tube amplifier where we easily find 70°C. This reduces the insulation resistance by a factor of 5. Still, even then, for MKT-caps the insulation resistance will remain as high as 200 GΩ; for our above example that would lead to an offset shift of 1 mV. That is a value that will not by any stretch of the imagination have an impact on the sound. Measuring various mostly old (but unused i.e. N.O.S. – new old stock) capacitors yielded values between 5 and 100 GΩ – that is clearly worse and could possibly classified as borderline regarding audible effects. However, really bad were the 0,15-µF-caps taken from an old VOX AC30 (of 1965 vintage): one still measured 2 GΩ but the other had dropped off to merely a 100 MΩ insulation resistance. The capacitance-values were still within the 20%-tolerance, but the leakage current shifted the operating point to an extent that in fact should have been designated a catastrophic failure. The amp, however, still worked, and whether the sound generated with this capacitor is judged as good or bad, as broken or as vintage, and the caps *therefore* are judged as junk or holy grails – that must be dealt with in the subjective domain.

Looking at things in a very fundamental way, it is possible that besides the purely electrical parameters, electro-mechanical parameters may also play a role. Indeed the coupling capacitor is charged via a high-impedance resistor, and if the capacitance changes over time, the capacitor acts as an AC-voltage-source – even without a guitar connected to the amp. The same principle as the one for a condenser microphone holds [3]: the high-impedance resistor (of e.g. 1 MΩ) prevents a quick charge transfer, and for an approximately constant charge, any small relative change in charge superimposed on top of this approximately constant charge corresponds to the change in voltage. Specifically: as the capacitance changes by 1‰, an AC-voltage of $U_{DC}/1000$ results. With a capacitor charged to 200 V, this would be 200 mV. Whether the capacitance can really vary by 1‰ is a different question. In a combo amp with speaker and amplifier in one and the same enclosure, we do find high sound pressure levels reaching 100 Pa and more. The resulting forces acting onto the capacitor housing will change the capacitance – but not normally by as much as 1‰. A simple consideration will help to estimate the order of magnitude: as a solid object is submerged in water, it is subject to a water pressure mounting with the submerge-depth. This pressure will crush even submarines made of steel if they dare to dive too deep – a capacitor however is much more fragile than a submarine. The higher the pressure, the more the capacitor electrodes will be pressed together. So, which dive depth might be equivalent to the above mentioned 100 Pa? Which special laboratory could be entrusted with finding this out? 100 Pa makes for 100 N per square meter … that corresponds to merely 1 cm dive depth! So: no special lab – the bathtub is good enough. Although: 200 V in the bathtub … no, better not. Dear music magazine journalists (if you at all accept advice from a scientist): do not try to do this at home! Danger to life! Only as a model experiment: the SPL generated in a combo is about as big as the water pressure at a depth of 1 cm. That should not deform a foil-capacitor to any substantial degree. For an orientating measurement, some brand-new 22-nF-capacitors were charged to 200 V and checked for microphonics: for SPL-values of 130 dB, the AC-voltage generated remained below 0,03 mV. Assuming 30 V to be actual ac-voltage at the plate, this microphonics-induced voltage would be smaller by factor of one million – for sure fully insignificant. Given the multitude of capacitor constructions that have found their way into guitar amps we cannot generally exclude that some capacitors would be among this crowd that exhibit much stronger microphonics – but the likelihood has to be seen as extremely small.

What remains to be looked at? How about reports such as: "*foil capacitors sound somewhat different than polypropylene*"? This statement opens up similar dimensions as "at night it is colder than outside". Very basically, capacitors may construction-wise be categorized into foil-, electrolytic-, sinter- and air-capacitors; for the dielectric, polystrol, polyester, or polycarbonate are in use, or the cited **polypropylene**. The typical polypropylene capacitor is a foil-capacitor consisting – in the KP build – of two foils on top of each other (metal and polypropylene, respectively), or – in the MKP build – of a metalized polypropylene foil. "If you hold two fingernails at a distance of one mm, you get a capacitance of 1 pF" – H.H. Meinke, unforgotten, r.i.p. To keep the in-between space in shape and enhance the insulation, we insert a thin foil in there, e.g. a foil of polypropylene. This also increases the capacitance by the relative permittivity (the relative dielectric constant) which for polypropylene amounts to about $\varepsilon_r = 2{,}2$, while for polyester it is 3,3. Both plastics belong to the group of **dielectrics** and therefore are insulating materials. The term "insulating" does however not imply that there are no charge carriers within them – the difference is that they are not as easily relocated. Current is nothing else than relocated charge: $I = \mathrm{d}Q / \mathrm{d}t$; i.e. no movement of charge, no current. In a copper wire the electrons can be very easily moved around (at an astonishingly low speed but in huge quantities), while in a dielectric there are next to no *freely movable* charge carriers present. Still, there are charges: positive atom cores, negative electrons, positive cations and negative anions. As a voltage is applied to the capacitor electrodes, forces are exerted onto the charge carriers, trying to shift and bend them; this is called the **polarization**. Since there are different kinds of charge carriers, there are also different kinds of polarization mechanisms. They are the cause of the **capacitor-losses**.

All materials are "built" of atom cores and atomic shells (model of the atom according to Bohr), and as an electric voltage is applied, an **electron-polarization** will occur in every material: the electric filed-strength shifts the electron shell relative to the atom core. This happens very quickly and is effective up into the THz region. In polar materials (e.g. polyester), the permanent molecular dipoles rotate under the influence of the external electrical field – this is called **orientation-polarization**. In materials containing ions, a counter-shifting of anions and cations occurs: this is the **ion-polarization**. Finally, it can happen in highly inhomogeneous materials that free charge carriers accumulate at insulating grain boundaries – here we have the **space-charge polarization**. All these polarization effects draw their actuation-energy from the electrical field and since none of these processes is reversible, part of the electrical energy is irreversible converted into heat. This caloric energy is not available to the electric circuit anymore (i.e. it is lost) – this is why we have "**losse**s". **Fig. 10.9.9** shows typical values.

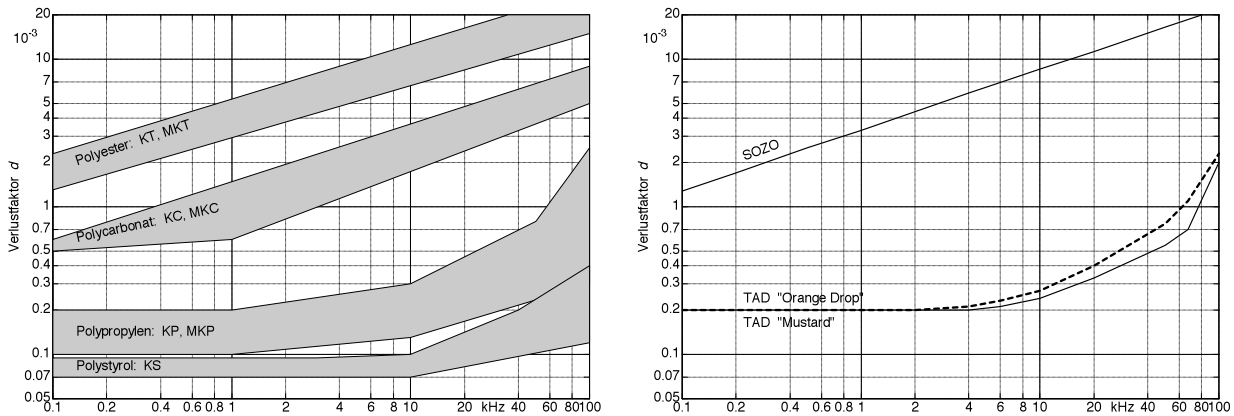
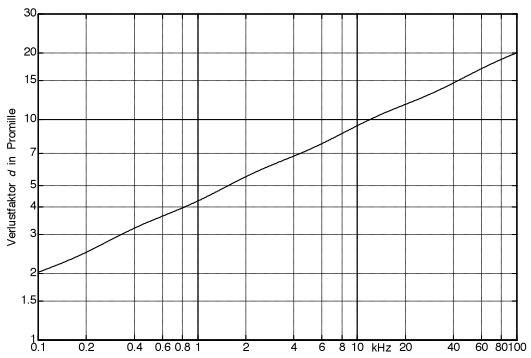


Fig. 10.9.9: loss-factors in typical coupling capacitors. Data-book information (left), measurements (right).

Borrowing from macroscopic effects, we could say: the microscopic polarization movements generate friction and corresponding losses; the latter are modeled as resistor(s) in the equivalent circuit. It was noted already at the beginning of this chapter that dielectric material properties are described 'merely' via ε and ρ. These material parameters are, however, frequency- and temperature-dependent (to bring up the most important influences), and in the general case also defined as direction-dependent tensors. Even using a simplified approach, conclusions based on the infinitesimal small cube and applied to the volume of the real capacitor do not result in merely *one single* resistor and *one single* capacitor but in a complicated network with actually an infinite number of components. It is, however, possible to recalculate this structure with good approximation into an impedance-equivalent (or impedance-like) circuit. This **equivalent circuit** has a big advantage over the capacitor model consisting of the two frequency-dependent components $R(f)$ and $C(f)$: it can be used to describe processes in time. The latter would be not easily handled with frequency-dependent components.



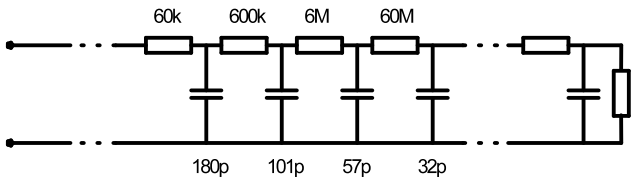


Fig. 10.9.10: Impedance-equivalent-circuit for a 22nF polyester capacitor (continued-fraction expansion).

Such an equivalent circuit is shown in **Fig. 10.9.10**. It is not the only possible one – depending on the desired accuracy, there are in fact myriad variants. In the diagram, we can see the slightly rippled approximation that could easily be improved at the expense of the number of components used. The chosen continued-fraction series expansion includes series-resistors the value of which rises, from left to right, by a factor of 10 each, and parallel-capacitors the value of which decreases, from left to right, by a factor of 1,78 each. (For a reduction of the ripple, both these factors need to be reduced). To the right, the "ladder" continues until we arrive at resistor values that correspond to the insulation resistance (lower cutoff frequency). The continuation to the left determines the high-frequency trend of the loss-factor. For the frequency range shown in the figure, the ladder does not need to be continued to the right at all if the given component values are used. To the left, the continuation needs to happen up to 60 Ω / 1 nF; a parallel capacitor (20 nF) and a series resistor (0,17 Ω) conclude the circuit. We can imagine that, for an insulation resistance of e.g. 1TΩ, the ladder is to be elongated further to the right, but it then becomes also clear how small the additionally included capacitances are (relative to 22 nF). For a capacitor terminated with extremely high impedance, this extension might be required, but for a typical tube circuit (1 MΩ), an equivalent circuit with a largest-resistor-value of 60 MΩ suffices as a good compromise.

Granted, this equivalent circuit is not that simple, either, but with today's computer-support, "impulse smearing" (group-delay distortion) can easily be determined. However, since a change in dielectric (e.g. to polypropylene) has no audible effect for the typical tube-coupling (as elaborated above at length), we will do without further explanations towards this.

Let us now turn to the question whether there could not be another reason why guitar magazines again and again report of **sound-changes** due to capacitors. Of course, we immediately remember the -20/+30% capacitance tolerances of the super-capacitors as mentioned above. However, first there are components boasting narrower tolerance ranges (the 60-€-mica-caps, for example, have a tolerance of *merely* 10%, at the most ☺), and second, the reports related mostly to changes in the *high*-frequency range. If we insinuate that this judgment does not relate to the MHz-range, but to the upper audio-range, what could be a technically supported reason? The size of the capacitors! Not the capacitance but the physical geometric dimensions. A coupling capacitor may come in axial build of the size ∅6x14, or the size ∅22x35 (mm each). Does size matter? Depending on circumstance, maybe yes. Since this capacitor is (relative to the resistors surrounding it) of low impedance at higher audio frequencies, the plate-ac-voltage is connected across it – independently of the capacitor's polarity. Between this electrode-surface of several square-centimeters and all conducting amplifier components, stray-capacitances result. In many guitar amps, the wire connected to the grid of the tube in question is of the un-shielded kind, and this will create a small capacitance between the coupling capacitor (plate) and the grid. This will not be a big capacitance, maybe 1 pF or 2 pF. Although every amplifier is put together a bit differently, with a big likelihood this capacitance will *increase* if a larger-volume-cap is incorporated. A mere 2 pF – that doesn't sound like much. However, we now need to consider the **Miller-effect** that increases (e.g. for an ECC83) the grid-input-capacitance by 100 pF (or even more) for any added 2-pF-grid-anode-capacitance. The tube itself has, according to the data sheet, C_{ga} = 1,6 pF, which yields (subject to the voltage gain) about C_E = 80 pF. Since the circuit build will not be totally free of capacitances, let us assume in the example C_E = 120 pF. This value would now be increased by the coupling capacitor to 220 pF. In conjunction with the source impedance we now arrive at a

low-pass with 7,2 kHz cutoff frequency.

Do compare this number with the Megahertzes cited in the capacitor adverts and do consider how big the reactance values could be here. Sure, not every amp has to be like that, indeed there are countless variants: Fender- and VOX-amps the insides of which deservedly have been called "birds-nests" of "cable jumble" already by other authors. Then there are boutique amps with wires bent at exactly 90° angles, fiber-boards, turret-boards, PC- and PTP-boards, source impedances of only 50 kΩ, but also of 250 kΩ, plus many more anomalies and peculiarities. And, indeed, stray-capacitances. So, as our guru introduces a hand-wound cap into the circuit with his heated iron while the circle of disciples holds devout silence, and as he calls for a listening test: maybe the sound of the amp has actually changed. This is because some wires were bent, because the plate-capacitors are moved closer to the grid wire by 1 cm, because the performing guitarist doesn't dare to dig into the strings as much in view of the horrendous price, or because the loss factor at 100 MHz has suddenly been reduced. There are even more possibilities, more things between heaven and earth, more knowledge, and more BS (not meant as abbreviation for Bachelor of Science). Science is not always welcome in this vicious cycle, and especially not the science of the electric current. Some authors in musician's magazines generally dismiss their perceived enemy ('*studied physicists'*) and advise to '*give scientists a wide berth*'. The latter will reciprocate right away, generally disqualifying every non-technician (or non-scientist ☺) as not having any ability to do scientific work.

Science requires reproducibility; the audiophile realm requires reproduction – that is something different. If an admirer of the arts buys a multi-million-€ painting not as an investment but just because he loves it, why would anybody carry out research to find a technical reasoning? Why would the aesthete want to know whether the green used by Gaugin might be greener by 0,221 nm than the green used by Dali? Somehow, this seems to be different in the area of audio-technology. Here, the Strat-player argues with the particularly high (or particularly low) weight of alder/ash, and the owner of a Plexi reasons with the particular group delay distortion of the Yellow Mustards. A 330B-fan will replace the polypropylene-capacitors by oil-paper caps because – as all the enlightened know – polypropylene is a synthetic, or, in lay-man's terms a plastic and so of course these caps will have that horrible, synthetic *plastic-sound* (advertisement). Do not ask whether an oily sound is actually preferable, because the 300B has entirely different problems: these oily comrades come either with aluminum foil, or with copper foil. Copper has better conductivity, and therefore – says the ad – the copper-sound will be better. A hand-wound aluminum cap will set you back 12 € a piece, but that is anyway more within the low-cost segment in these circles, and does not really match the matched triode-pair (at 250 €). And so copper-foil it is, because: the conductivity is 60% better, and the price is 100% higher – that works as a beginners-set. For the next birthday, we will nevertheless rather reward ourselves with the real deal: with **silver-foil capacitors**, because: silver has still better conductivity, says the ad, and who but the webpage of the manufacturer would know better. So: silver. There's a lingering memory from that dreaded latin class: silver – argentum – Argentarius? Sin-offering … no: money business! That fits: big money business, because: there's not just one coupling capacitor in that radio – er: guitar amp, but there are two … no: three. Per channel! O.K., there's the little box on the on-line order-form: enter "6". And stay strong, as in the box on the bottom the sum appears: 1101,00 €. 'Tis the birthday – off into the shopping basket, done. Well … just to be safe, enter "resubmission" for the next but one birthday – at the latest, replacements should be acquired then, because: for Ag-caps, the manufacturer explicitly mentions the minimum life-time: 2 years. That's not difficult: acquire, solder in, wait for the burn-in time to pass, listen, buy replacements, solder in, wait for the burn-in time to pass, and so on. And in case anybody has any doubts at all about these mod(ification)s: data tables from electrical engineering: indeed, the conductivity of silver is better than that of copper by 6%. *Though this be mod-ness yet there is method*, or so Shakespeare notes.

The capacitance-tolerance of these money-capacity-robbing darlings is specified to +30% … o.k., it is what it is, don't get wound up, they are hand-wound. "*Quality has its price (sic)*" You should not take too narrow a view on the fact that the auditory system can muster the cited µs-resolution – if at all – only inter-aurally i.e. "between the channels". The audiophile writes in an internet chat room: *hopefully this tolerance will not have a big impact in front of a tweeter?* No, no worries – tweeters are generally known to be very tolerant towards minorities. Plus, if indeed any uneasiness remains, for sure there will be someone offering – for something like 2022,00 € – a selected version with smaller tolerance. Don't you even think about the 1%-filter-caps! They are down the cheap end, and there's no way they can sound at all. If only the best is good enough: **selected Ag-caps**. Grab them every other year, or every 10 000 km, whatever comes first.

By the way, what would the synthesis of idiographic♠ und diotic♠ be? Audiophile??

♠ idiographic = describing the very special; diotic = listening with the same signal at both ears.

Dielectrics for capacitors

Mica
Up to 125°C (max. 155°C). Relative permittivity $\varepsilon_r = 5{,}5 \ldots 7$.
Also: Phlogopite mica, Micanite, Micalex, mica foil, Samikanite (with different data).
Mostly electron- and ion-polarization. Losses are frequency-independent (≈GHz).
Highest stability of capacitance over time; smallest temperature-coefficient.

Polystyrene (**KS**, Styroflex)
Thermoplastic, mostly electron-polarization.
Since 1936 up to 60°C, since 1953 up to 70°C (max. 85°C). Relative permittivity $\varepsilon_r = 2{,}5$.
Very high insulation resistance, very small losses.

Polypropylene (KP)
Thermoplastic, mostly electron-polarization.
Available since 1960; up to 85°C. Relative permittivity $\varepsilon_r = 2{,}3$.
Very high insulation resistance, very small losses.

Polycarbonate (KC)
Thermoplastic, mostly electron- and orientation-polarization.
Available since 1961; up to 100°C, max. 125°C. Relative permittivity $\varepsilon_r = 2{,}8 \ldots 3$.
Very high insulation resistance, very small losses.

Polyethylene terephtalate (**KT**, Polyester)
Thermoplastic, mostly electron- and orientation-polarization.
Available since 1957; up to 100°C, max. 125°C. Relative permittivity $\varepsilon_r = 3{,}3$.
High insulation resistance, small losses.

Paper, impregnated (P, MP)
Sulfate cellulose, mostly electron- and orientation-polarization.
Characteristics depend strongly on density, water content and impurities.
Depending on the situation only moderate insulation resistance, small losses. Max. 100°C.

Capacitor-oil (Naphthenic oil etc.)
mostly electron-polarization; however: oxidization products (acids) are polar.
Relative permittivity $\varepsilon_r = 2{,}2$. Copper will accelerate the oxidization of the oil.
Depending on the situation very limited life-time.

Al_2O_3, Ta_2O_5
For electrolytic capacitors, not used in coupling capacitors.

Ceramics, e.g. TiO_2
For ceramic capacitors; not used in coupling capacitors.

10.9.4 Sound event vs. auditory event

On the one hand, it is possible to document the operational behavior of a guitar amplifier via formula and results of measurements; on the other hand, it may happen by verbal description of sensory perceptions. "*Smells like a goat*" would be a genre-typical choice of words, or "*has one hell of an oomph and creates just the right sizzle*", so stick with auditory perception. If everybody knows what an oomph is, this description indeed does help. However, because scientists often do not know what an oomph is, and because they like to quantify things into interval- and relational scales, there are also numerical specifications such as "*cutoff frequency at 5238 Hz*". So, we have, on one side, physics with its objective sound-event data: 100 W, 8 Ω, 5238 Hz, 10 ms. On the other side we find the auditory event with verbal, subjective judgments; louder, much more authentic, vintage-like, throaty sound, too short sustain, etc. In between there is the magnitude estimation: twice as loud as … , just noticeable reverb amount, 50% longer sustain.

Guitar amps mostly do not play for measuring equipment but for people. Okay, they also play for tables, chairs, the dogs of innkeepers and their fleas, but predominantly for people, after all. Whether a measuring device certifies an increase of the effects-mix from 1% to 2% is insignificant if this remains inaudible in both cases. The physical sound event leads – if it is audible – to an **auditory event**, and it is only the latter that is judged by the listeners. The assessment is anything but objective: whether an amp-sound is judged as being good or bad is a matter of taste and depends on subjective criteria and also on environmental conditions. Everybody knows **optical illusions,** and there is no surprise in the fact that there may also be auditory illusions. Nobody will assume that a car speeding away on a straight road actually decreases in size although the optical angle that it occupies in our visual perception indeed becomes smaller. The brain will correct for the shrinking image on the retina and, in a way, creates an illusion. Is it actually an illusion? The car has not shrunk, after all, just the picture on the retina! Anyway, the term "optical illusion" found its way into everyday language.

What is the reason for such illusions? Is a lion that only then a lion when we see it in full, or is it a lion already as it steps out of the bushes, only half visible? This is a clear-cut case of evolution and/or selection. It was conducive to survival to supplement fragmentarily arriving perceptions, and to correct distorted sensory impressions. The immense flood of data arriving from our sensory receptors needs to be reduced momentarily by many orders of magnitude: the data flow taken from a stereo CD amounts to about 1,4 Mbit/s but at best only 50 bit/s of that arrives at our consciousness. However, the synapses working on our internal signal-processing do not just throw, without discretion, 99,996% of the incoming information into the bin; there are rules – but rules that may change from one second to the next, with our cooperation but also without. Since we perceive our environment exclusively through this information-reducing filter, the philosopher arrives at the conclusion: **nothing is as it seems** – and he seems to be right. The "seems" is attributed to the realm of the perceptions (auditory event), the "is" to the realm of physics (sound event). It must not surprise us if a guitarist perceives sound changes if he is being told that a coupling capacitor has been swapped – although the amp remained in fact untouched, and merely the judgment criteria have undergone a change. The opposite may also happen: a capacitor is indeed swapped but nobody hears a difference. And of course there is the third variant: the swap is clearly audible. There are countless guitar amps, if not more – for the individual case no remote diagnosis can be established. The following explanations can therefore only impart basic knowledge but not offer retrofitting plans for specific amplifiers.

Fig. 10.9.11 shows some optical objects. In the first picture we see two crossing straight lines, in the second two overlapping circles. Or are these in fact other objects? Aren't there two angles with meeting apex in the first picture? We could just as well assume that – but the crossing straight lines are simply more obvious. **Our brain always chooses the interpretation of reality that is more likely**. In this case, this is the crossing of two lines (or two tree branches that have fallen on top of each other). For the same reason we do not recognize, in the second picture, a crescent and a waning moon with a convex lens-type area in between, but two circles. In the third picture, we see two triangles on top of each other that do not at all exist in the drawing. In particular, the "upper", white triangle is predominantly "make-believe" rather than "actually being". The right-hand picture conveys a depth in space that is not at all present in reality. And although this picture does not change, it can "jump" in our perception: one moment we see a cube on the floor, the next we see a cube hanging (fastened with its rear surface to wall) towards the left … or towards the right. Visual perceptions seem not to correlate perfectly with the optical stimuli.

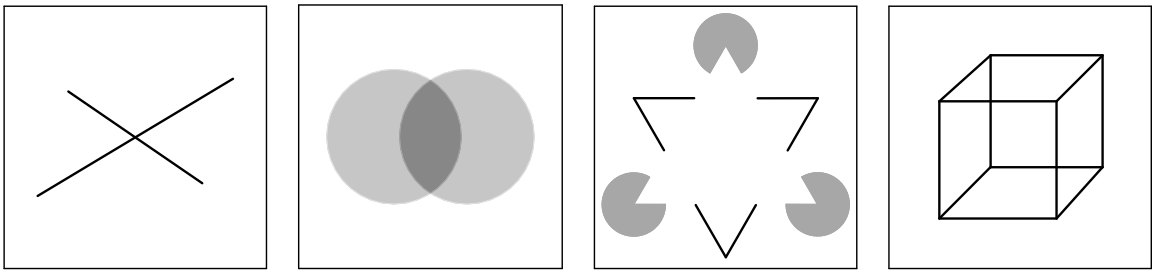


Fig. 10.9.11: Examples regarding the visual perceptions of optical objects. For more examples see D. Picon 2005.

Consequently, we should not be puzzled if auditory perceptions change as well, without any alteration in the acoustical sound event. A special experimental methodology is necessary to establish whether or not there is in fact a **causal correspondence** between a change in our auditory perception and a change in the physical sound event. How would a guitarist who has just swapped a capacitor in his amp (and now plays to check out the result) judge whether any perceived difference in sound is due to the changed capacitor, or due to the (unconsciously) changed way of playing, or due to the (unintended) change in the listening position, or due to changed judgment standards (autosuggestion)? Psychometrics has a few hints here: for example, the sounds to be judged should be presented such that the test person does not know which sound is presented at the given time ("blind"-test). The sounds should have a duration of only a few seconds, and the interval between sounds should be short (about 0,5 s). In a comparison of pairs (A-B-A-B) only a single parameter should be changed at a time. How much does a demo-CD for replacement pickups tell us if there is a different guitar-riff for each pickup, and if possibly different players have recorded the riffs? Not much!!

The first run-through of a listening experiment could, for example, contain simple **nominal verdict**s: the perceived sounds sound **the same or different**. To increase the certainty of the statements, it is necessary to have the subject judge identical sounds without the subject knowing that this pairing is included. A subject that repeatedly hears differences when identical sounds are presented (perceived A-B-A-B is in reality A-A-A-A) will either uncover faults in the experimental setup, or he/she is unsuitable as a test subject. If two sounds are, objectively, not significantly distinguishable, the question about which sounds better is moot.

If A and B are judged as sounding different in the auditory experiment, the second experimental stage can serve to ask about comparative ranking characteristics (**ordinal characteristics**): "*I like B better than A.*", or "*B sounds more distorted than A*", or certainly even "*B has more oomph that A*". In the last stage♣, quantitative **cardinal characteristics** are addressed: "*I would spend 100 € more for B.*"

In order to judge the subjective difference in sounds between A and B, the exact objective difference between these sound needs to be known – that should be a matter of course. For a listening test on the sound of capacitors (Chapter 10.9.3), this implies that the amplifier is always driven by the same signal, i.e. not by a guitarist (with a guitar) playing this now and that then. Rather, the guitar is recorded *once* in an appropriate way, and this recording is fed to the amp in an identical manner for the listening test. Specialist knowledge is indeed required in order not to destroy the sound already by the experimental setup. As a result, the following could be obtained: "*Of 20 subjects only 3 could hear a difference between A and B.*" Or something like: "*15 of 20 subjects judge A as sounding better but would on average accept no more than 10 € additional cost.*" could be the result. Still, even such tests leave questions unanswered: anybody who has not personally participated will not know whether he/ she would belong to a) the 15 or to b) the remaining 5, and if a), then the pecuniary equivalent might be as much as 500 €, as well. In general: if I am asking for the opinion of someone else, then I will receive the opinion of someone else – that is highly trivial. If I want to rely solely on my own opinion, then I need to test everything myself (and why not?). If I do ask another person, I might be i.a. interested in how reliable this person's opinion is. In such a case this approach holds: for a prejudice-free subjective judgment of objective issues, blind tests provide a powerful tool.

But what about those instances when the sound of an amp changes without identifiable objective reason? Those cases when an amp has lost its unique sound after a repair job, although it was – embarrassment city! – accidentally shipped back without having been opened up? The case of the guitar that never sounded right again after it had been kidnapped for a stage-quickie by a pal. Or the case of the capacitor-swap that led to a sound miracle although everybody (or rather all "studied physicists") tirelessly continues to emphasize this to be impossibly? There could be physical reasons (transport, shift of a slightly loose guitar neck, stray capacitances), but we might also see in such cases the impact of judgment benchmarks that are easily influenced. Most people fancy themselves to be superior to the average in many areas, and prefer that their equipment to stand out from the mainstream: alloy wheels … or copper caps. No sooner than a prejudice takes hold, it is pampered and cultivated – the smallest confirming hint is scraped up and blown out of proportion while every counterargument is conveniently ignored. As a rule, every confirmation is trustworthy while every disagreement is questionable. No one is spared this kind of delusion: 94% of all scientists at university deem their research to be above average! *The deeper reason for our biased dealings with information stems from a conflict between the search for truth, and the search for harmony and for agreement with ourselves. To admit that one has been wrong can, after all, chip away at one's self-esteem and one's image.* [R. Degen, Lexikon der Psycho-Irrtümer – *lexicon of psycho-errors*]. This is why an assumed change can lead to a change in perception. If, after 100 h of playing the new capacitors, suddenly the treble comes to life, the underlying mechanism is not necessarily an objective reason – the belief is already sufficient. It is a rather big paradox that training can render our hearing more precise but at the same time more susceptible to influence.

♣ These results may be achieved as well in a single run-through, if a matching evaluation-statistic is employed.

That the brain can be trained is without a doubt. **Practicing** for many years fine-tunes the auditory performance, makes small differences stand out, allows for more comparison patterns to be available, and enlarges the sensory areas in the cortex. From the awareness of above-average hearing-prowess, the idea can easily arise that "the whole hearing" is now perfected and has become the unswayable calibration-standard. In this, it is easily overlooked that numerous auditory functions are not (or only to a very small degree) trainable, after all – they function just as they do for the untrained and are therefore – relatively seen – worse off than at the beginning of the training process.

An example from optical processing: for the cube in Fig. 10.9.11, we can decide whether we want to see it as one whole object (the cube), or as individual lines. Everybody with normal vision can do that; it does not require special training. For acoustical objects, however, different rules apply: in a complex sound made up from partials (harmonics) it is much harder to hear individual partials; often it is even entirely impossible. A simple trick may help: a special (non-masked) partial is switched off (filtered out) for a short time and then switched on again. At the switching-off instant we hear, as expected, a change in sound (thinner, more hollow). As the partial is switched back on, there is a surprising effect: first, the thinner, more hollow remaining sound is joined by an individually audible sine-tone that "melts" into the remaining sound within a few seconds to eventually form the original sound. Something new, especially when appearing abruptly, is deemed important, and the brains switches to "make individual object audible"-mode. After some seconds, the new additional object is categorized as a kind of prodigal son perfectly fitting in with all other objects, and the precedence circuit is switched off again: the partial is not audible per se anymore. No training can change this effect. The auditory perception changes although the sound remains static! On top of such autonomous (endogenous) signal-processing algorithms, other external (exogenous) signals affect the perception process: directional hearing is influenced by visual clues, as well, as is the impression of reverberation and even speech intelligibility. Nothing is, as it appears, and everything appears different

A real-life example shows how difficult listening test can be: in a pretty hefty pickup comparison test (Gitarre&Bass 2/05), there are 10 pages of verbal assessments: *"In comparison almost mushy … the picking attack substantially softer and brittle … surprisingly glassy and rich in harmonics … an entirely different spectrum in the mids … far less richly colored … acutely transparent and translucent … a sound beautifully soft and compressed … a very creamy tone that however seems a bit dull and lackluster … although completely covered in wax, the pickups sound open and as airy as un-potted ones.*" These short excerpts indicated that clearly audible differences must exist between the judged pickups. Some 2 years later, the same magazine publishes a flash-back to the same test. This flash-back arrives at the conclusion that "*in fact all models sounded almost the same."* (Gitarre&Bass 5/07) The difference between "*entirely different*" and "*almost the same*" has to be seen – according to the flash-back – in the different recording situation. Mind you: for "*almost the same*", the recordings were not done in a garage but again in the recording studio, and getting "*good and professional results.*" Based on this, every reader can pamper his personal prejudice: one will shell out 400 € for a pair of PAF-clones and enjoy the exclusivity, the other will (because of "*almost the same*") stick with the equipment he already owns, and prefer to perfect his finger vibrato – chacun à son goût. Another one may comment on the published sound examples from the above test with "*You must all be mad! There's nothing to hear but one and the same pickup again and again!*" (Gitarre&Bass 4/08). It does dignify the author of the article that he has not withheld this comment from his readers.

