Table of Content

1 Overview

This amplifier is a standard three-stage AB amplifier and is inspired by Douglas Self's amplifier designs.

The reason why I made this documentation is the missing calculation on the Internet for beginners. On the Internet, only fractions of circuits are calculated or Douglas Self doesn't publish his formulas. This cookbook should help beginners and does not deal with detailed descriptions or derivations. The fundamental theory of the operation of differential amplifiers, voltage amplifiers, constant current sources and class B, AB amplifiers should already be known.

I don't guarantee the correctness of the following calculations. However, my self-built amplifier works blameless.

My AB-Amplifier can be found here: http://members.mvnet.at/sikado/

This homepage is written in German because I am from Austria. Furthermore, it is not professional and the only purpose is to show my DIY projects.

Please contact me under si1505do@gmail.com if you find a mistake or you have some questions.

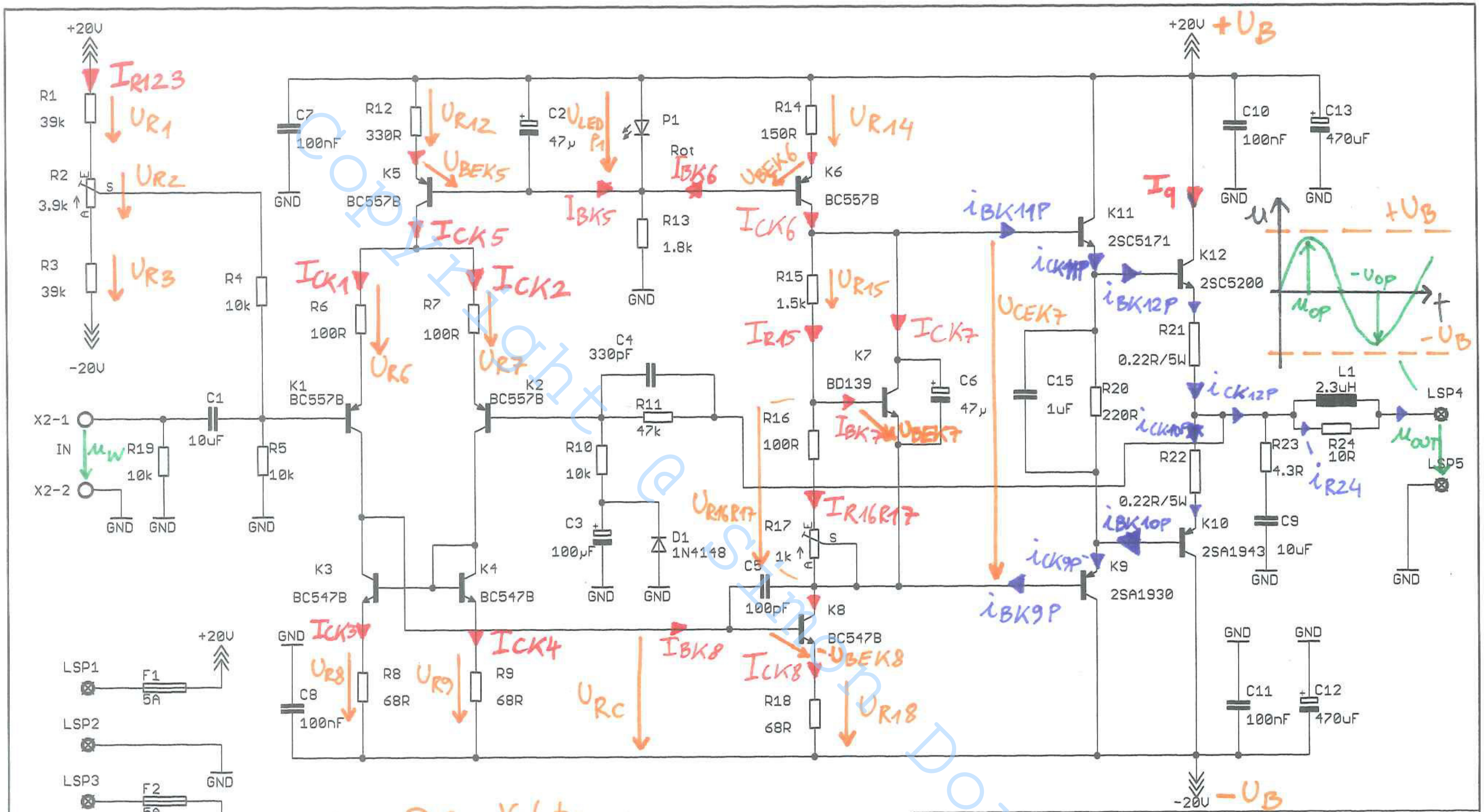
The circuit of this calculation can be found in the attached .pdf file on the last pages.

In the calculations, U and I are DC, u and i are AC!

2 Circuit with all calculated voltages and currents

Here, the circuit is attached with all calculated voltages and currents to understand all used variables.

Keep in mind, that R_{20} and C_{15} are wrong in this circuit. This circuit is an old version and only serves to illustrate the current and voltage variables. The latest and stable version is attached at the end of the document.



■ ... DC - Voltage

■ ... AC - Voltage

■ ... DC - Current

■ ... AC - Current

TITLE: Endstufe_30W

Document Number:

REV:

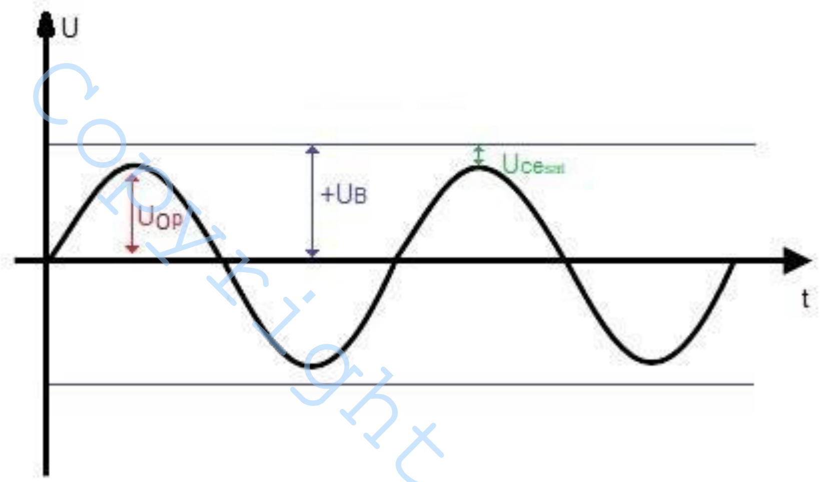
Date: 07/10/2018 21:57:52

Sheet: 1/1

3 Class AB output stage

3.1 Output sinewave:

First, we should have a closer look at the output sine wave. The maximal output peak voltage (u_{OP}) is the power supply voltage ($+U_B$) minus the transistor collector-emitter saturation voltage (U_{CEsat}). The smaller U_{CEsat}, the higher the output power.



This amplifier is supplied with a toroidal transformer of 2x15VAC. The minimal VA for this transformer is calculated later. 2SA5200 transistors are used, which have a saturation voltage of $U_{CEsat} = 0.2V$. The minimum load resistance is $R_L = 4\Omega$.

given.: $+u_{AC} = 15V$, $U_{CEsat} = 0.2V$, $R_L = 4\Omega$;

3.2 Supply voltage ($+U_B$):

The calculation of the power supply is kept short. For more details google "full bridge rectifier", tons of formulas can be found.

For the power supply, an integrated full bridge rectifier and capacitors are used. The forward voltage of one diode is $U_D = 0.55V$.

3 x 15000uF capacitors are used in parallel to reduce the voltage ripple. Therefore, the hum voltage is really small and can be neglected.

$$+\mathbf{U_B} = u_{AC} * \sqrt{2} - 2 * U_D - u_{hum} \approx u_{AC} * \sqrt{2} - 2 * U_D = 21.21V - 1.1V = \mathbf{20.11V}$$

For simplification, $+U_B$ is assumed with 20V. $-U_B$ equals $+U_B$ because of the symmetrical power supply.

3.3 Amplifier output peak voltage U_{OP}:

As described above, u_{OP} is $+U_B$ minus U_{CEsat}.

$$\mathbf{u_{OP}} = +U_B - U_{CEsat} = 20V - 0.2V = \mathbf{19.8V}$$

U_{CEsat} is assumed with 0.2V at the beginning. The final result can be found under Measurements.

3.4 Emitter resistors R_{21}, R_{22}:

These resistors are for DC operating point stabilization and reduce AB crossover distortion. The current gain (β) of the push-pull transistors are not the same. R_{21} and R_{22} should compensate this difference.

A rule of thumb for these resistors is:

$$\mathbf{R_{21} = R_{22} < 0.1 * R_L \approx 0.22\Omega}$$

With R_L of 4Ω, resistors smaller than 0.4Ω should be taken. For smaller output power (<100W), 0.22Ω is suitable. For higher output power (>100W), 0.1Ω should be used, otherwise, the power loss at the resistors is getting too high.

3.5 Peak and RMS output power:

P_{OP} is the output power with the transistor power loss and the emitter resistor power loss.

The peak output power (P_{OP}) equals:

$$\mathbf{P_{OP}} = \frac{1}{2} * \frac{u_{OP}^2}{(R_L + R_{21})} = \frac{1}{2} * \frac{(19.8V)^2}{(4\Omega + 0.22\Omega)} = \mathbf{46.45W_{PEAK}}$$

The RMS output power (P_{ORMS}) equals:

$$\mathbf{P_{ORMS}} = \frac{P_{OP}}{\sqrt{2}} = \mathbf{32.85W_{RMS}}$$

3.6 Supply power / transformer power:

$$\mathbf{P_S} = \frac{2 * +U_B * u_{OP}}{\pi * (R_{21} + R_L)} = \frac{2 * 20V * 19.8V}{3.14 * 4.22\Omega} = \mathbf{59.77W}$$

The 2 in the formula expresses the symmetrical power supply. For a symmetrical power supply for +-20V, a transformer with 2x15VAC and at least 60VA should be used. Keep in mind, that the VAS and the differential amplifier also need some current.

3.7 Peak power loss of the emitter resistor:

The peak resistor power loss (P_{R21P}) equals:

$$\mathbf{P_{R21P}} = \mathbf{P_{R22P}} = P_{OP} * \frac{R_{21}}{(R_{21} + R_L)} = 46.45W_{PEAK} * \frac{0.22\Omega}{4.22\Omega} = \mathbf{2.42W_{PEAK}}$$

The RMS transistor power loss (P_{R21RMS}) equals:

$$\mathbf{P_{R21RMS}} = \mathbf{P_{R22RMS}} = \frac{P_{R21P}}{\sqrt{2}} = \mathbf{1.71W_{RMS}}$$

It can be seen, that the power loss of the resistor is getting higher with increasing resistance.

3.8 Power loss of the transistor:

The supply power minus the output power and the resistor power loses is the power loss of the AB transistors.

The peak transistor power loss (P_{K12P}) equals:

$$\mathbf{P_{K12P}} = \mathbf{P_{K10P}} = \frac{1}{2} * (P_s - P_{OP} - P_{R21P}) = \frac{1}{2} * (59.77W - 46.45W - 2.42W) = \mathbf{5.45W}$$

The RMS resistor power loss (P_{K12RMS}) equals:

$$\mathbf{P_{K12RMS}} = \mathbf{P_{K10RMS}} = \frac{P_{K12P}}{\sqrt{2}} = \mathbf{3.85W}$$

The power loss of the transistors does not reach its maximum at the maximum voltage level (u_{OP}). The maximum power loss is reached at:

$$\mathbf{u_O} = +U_B * \frac{2}{\pi} = 20V * 0.64 = \mathbf{12.8V}$$

$$\mathbf{P_{K12max}} = \mathbf{P_{K10max}} = \frac{1}{\pi^2} * \frac{+U_B^2}{(R_L + R_{21})} = 0.1 * \frac{(20V)^2}{4.22\Omega} = \mathbf{9.48W}$$

This high amount of power loss needs an appropriate heatsink. This calculation can be found under Heatsink calculation.

The lower the saturation voltage, the lower the power loss of one transistor. The saturation voltage increases with lower collector current. So, the best power efficiency is reached at the maximum voltage level.

3.9 Power efficiency:

$$\boldsymbol{\eta} = \frac{P_{OP}}{P_S} = \frac{46.45W}{59.77W} = 0.78 = \mathbf{78\%}$$

η describes the power efficiency without the quiescent current (Class B Amplifier). That is why, the power efficiency of a class AB amplifier is smaller than 78%, depending on the set quiescent current. The final power efficiency can be found under Measurements.

3.10 Output transistor reverse voltage:

In the push-pull technology, only one transistor is on, the other one is blocking the negative voltage. Therefore, the chosen transistors should have a high reverse voltage.

A rule of thumb for the reverse voltage is:

$$\mathbf{U_{rev}} = \mathbf{V_{CEO}} = 3 * +U_B = 3 * 20V = \mathbf{60V}$$

3 * are selected for safety, as inductive load (speakers) can cause induced voltage spikes. The reverse voltage is V_{CEO} in the datasheet. The 2SC5200 has 230V V_{CEO}.

3.11 Collector current of the output transistors:

The peak collector current equals:

$$\mathbf{i_{CK12P}} = \mathbf{i_{CK10P}} = \frac{u_{OP}}{(R_L + R_{21})} = \frac{19.8V}{4.22\Omega} = \mathbf{4.69A}$$

The arithmetical average current equals:

$$\mathbf{i_{CK12\pi}} = \mathbf{i_{CK10\pi}} = \frac{i_{CK12P}}{\pi} = \frac{4.69A}{3.14} = \mathbf{1.49A}$$

3.12 Base current for output transistors:

<u>Hint:</u> You should always take the minimum current gain from the datasheet. If the current gain is higher in praxis, no problems can occur compared to the maximal current gain (less current).

$$\boldsymbol{\beta} = \boldsymbol{\beta_{K12}} = \boldsymbol{\beta_{K10}} = \mathbf{50}$$

$$\mathbf{i_{BK12P}} = \mathbf{i_{BK10P}} = \frac{i_{CK12P}}{\beta_{K12}} = \frac{4.69A}{50} = \mathbf{93.8mA}$$

The maximum peak base current equals 93.8mA.

3.13 Quiescent current:

A rule of thumb expresses that about 10mV should drop across R_{21} and R_{22} in quiescent mode. The higher the quiescent current, the worse the power efficiency. But if the quiescent current is too low, crossover distortion can occur. This current is set with the potentiometer (R_{17}) of the U_{BE}-Multiplier.

From this rule follows:

$$\mathbf{I_q} = \frac{U_{R21}}{R_{21}} = \frac{0.01V}{0.22\Omega} = \mathbf{45mA}$$

The quiescent current is finally set correctly with the oscilloscope. 45mA is only a guideline.

3.14 Gain of the AB-Output stage:

- The current gain (V_{iOS}) equals:

$$\mathbf{V_{iAB} = \beta}$$

- The voltage gain (V_{UOS}) equals:

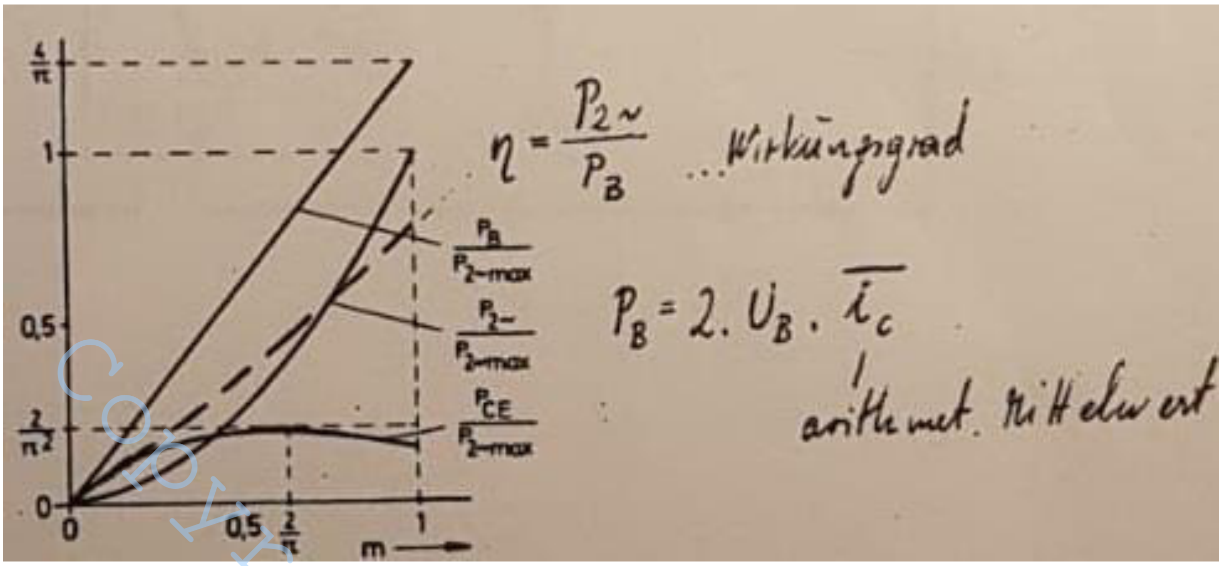
$$\mathbf{V_{UAB} < 1 \approx 1}$$

3.15 Output resistance of the AB-Amplifier:

$$\mathbf{r_{OUT}} = R_{21}||R_{22} = \frac{R_{21} * R_{22}}{R_{21} + R_{22}} = \frac{(0.22\Omega)^2}{0.44\Omega} = \mathbf{0.11}\Omega$$

It can be seen, that the output stage has a really low output resistance. That is why, an AB-Amplifier works as a voltage source.

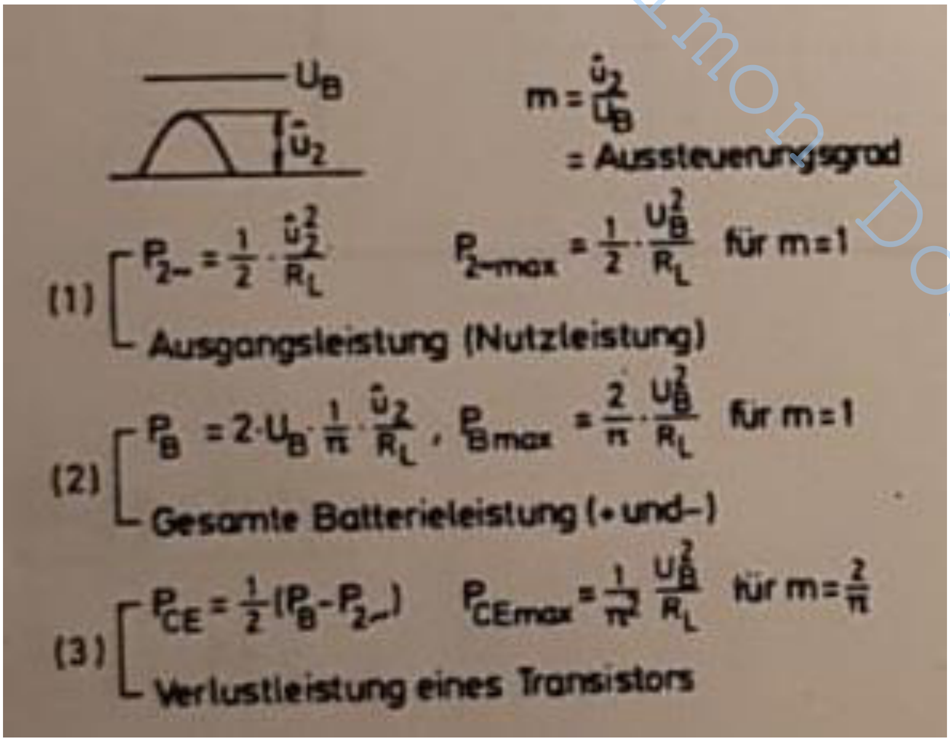
3.16 Power diagram:



In this diagram, the different power losses depending on the ratio of maximal output voltage level and supply voltage (m) can be seen. At $\mathrm{m} = \frac{2}{\pi}$ the maximal transistor power loss occurs. This diagram verifies the above-calculated power losses.

3.17 Summary of formulas:

These formulas are from my German-language lecture at the higher technical college.



4 Output driver stage

Mostly, power transistors have a low current gain. So, at high output currents, the base current is also relatively high. Often, this base current cannot be supplied by the VAS. Thus, for the base current of the output transistor, another driver stage is needed.

The VAS has to deliver 93.8mA of current without a driver output stage!

4.1 Base current for driver transistors:

$$\mathbf{i_{CK9P}} = \mathbf{i_{CK11P}} = \mathbf{i_{BK12P}} = \mathbf{i_{BK10P}}$$

$$\beta = \beta_{K9} = \beta_{K11} = 100$$

$$\mathbf{i_{BK9P}} = \mathbf{i_{BK11P}} = \frac{i_{BK12P}}{\beta_{K11}} = \frac{93.8\text{mA}}{100} = 0.94\text{mA} \approx \mathbf{1mA}$$

It can be seen, that this current is one hundredth smaller than the base current for the output transistors. This AC current of 1mA can be supplied easily from the VAS.

4.2 Resistor R_{20} and Capacitor C_{15} / Resistors R_{20} and R_{25}:

Switch-off distortion is a phenomenon that occurs when a transistor has its base driven from a high impedance source and is suddenly pushed to a lower base voltage. In the ideal case, the transistor would react immediately to this change. However, there will be some charge left over in the base region of the transistor that has nowhere to go except into the transistor channel and through the emitter. The consequence is that the transistor keeps on conducting for a while even after being switched off. The remedy for this is to provide an easy discharge path for the base charge. This can be seen in the old schematic in the form of the resistor R_{20} and the capacitor C_{15}. Unfortunately, some changes had to be made because the power amp was unstable. These changes can be found in the new schematic in the form of the resistors R_{20} and R_{25}.

Unfortunately, I didn't found any formulas for this two components. That is why these two values are taken directly from the Douglas Self amplifier.

$$\mathbf{R_{20}} = \mathbf{R_{25}} \approx \mathbf{47\Omega} \ldots \textbf{for smaller output power}$$

$$\mathbf{R_{20}} = \mathbf{R_{25}} \approx \mathbf{22\Omega} \ldots \textbf{for bigger output power}$$

4.3 Driver transistor reverse voltage:

In the driver stage, only one transistor is conducting and must block the other voltage. Therefore, the driver transistor also needs 3 x $+U_B$ as reverse voltage.

$$\mathbf{U_{rev}} = \mathbf{V_{CEO}} = 3 * +U_B = 3 * 20V = \mathbf{60V}$$

The 2SC5171 has a V_{CEO} of 180V.

4.4 Gain of the output driver stage:

- The current gain (V_{iDS}) equals:

$$\mathbf{V_{iDS}} = \boldsymbol{\beta}$$

- The voltage gain (V_{UDS}) equals:

$$\mathbf{V_{UDS}} < \mathbf{1} \approx \mathbf{1}$$

5 U_{BE} – Multiplier

This circuit is used to produce the biasing voltage of 4 * 0.7V = 2.8V for all four base-emitter paths. With this voltage, every transistor is conductive and the crossover distortion can be reduced to its minimum. The higher the biasing voltage, the higher the output quiescent current. So, the potentiometer has to be calculated carefully.

5.1 Constant current source for U_{BE}-Multiplier and VAS:

The constant current source should deliver 10mA and is stabilized by a red LED. The red LED has less noise than a common Z-Diode. However, the remaining LED noise is filtered with a 47uF Capacitor (C_2). Furthermore, C_2 should boost the constant current source, when the supply voltage is dropping during bass drops.

A small current of the LED current is also the base current of K_5 and K_6 (I_{BK5K6}). This current is very small (uA) and can be neglected.

given: $+U_B = 20V$, $U_{LEDP1} = 2.1V$, $U_{BEK6} = 0.7V$, $I_{CK6} = 10mA$, $\beta_{K6} \approx 300$, $I_{LEDP1} = 10mA$;

- R_{14} – Calculation:

$$I_{CK6} = \frac{U_{LEDP1} - U_{BEK6}}{R_{14}} \rightarrow \mathbf{R_{14}} = \frac{U_{LEDP1} - U_{BEK6}}{I_{CK6}} = \frac{1.4V}{0.01A} = 140\Omega \approx \mathbf{150\Omega} \rightarrow \mathbf{I_{CK6} = 9.3mA}$$

The current through the K_8 transistor is the constant current minus the quiescent current of the driver transistor. The quiescent current on the output is 45mA, which equals 9uA base current for K_9 and K_{11}($I_{BK9q} = I_{BK11q}$). This current is very small and can be neglected. The AC current i_{BK9P} and i_{BK11P} is delivered by the VAS and has no impact on the quiescent calculations.

$$\mathbf{I_{CK8}} = I_{CK6} - I_{BK9q} - I_{BK11q} \approx I_{CK6} = \mathbf{9.3mA}$$

- R13 – Calculation:

$$\mathbf{R_{13}} = \frac{+U_B - U_{LEDP1}}{I_{LEDP1} + 2 * I_{BK5K6}} \approx \frac{+U_B - U_{LEDP1}}{I_{LEDP1}} = \frac{21V - 2.1V}{10mA} = \frac{18.9V}{0.01A} = 1890\Omega \approx \mathbf{1.8k\Omega}$$

- R_{iK6} – Calculation (K_5):

$$\mathbf{R_{iK6}} \approx r_{CE} * \frac{\beta_{K6}}{1 + \frac{\beta_{K6} * U_T}{U_{R14}}} = \frac{100V}{I_{CK6}} * \frac{\beta_{K6}}{1 + \frac{\beta_{K6} * U_T}{U_{R14}}} = \frac{100V}{9.3mA} * \frac{300}{1 + \frac{300 * 25mV}{1.4V}} = \mathbf{507k\Omega}$$

5.2 Calculate U_{BE} – Multiplier:

$$\mathbf{I_{CK7} = I_{CK8} = 9.3mA}$$

$$\mathbf{I_{BK7}} = \frac{\mathbf{I_{CK7}}}{\beta_{K7}} = \frac{9.3mA}{40} = \mathbf{233uA}$$

$$\mathbf{I_{R16R17}} = 5 * I_{BK7} = 5 * 233uA = \mathbf{1.16mA}$$

$$\mathbf{I_{R15}} = I_{BK7} + I_{R16R17} = 233uA + 1.16mA = \mathbf{1.4mA}$$

$$U_{CEK7} = U_{BEK7} * \left(1 + \frac{R_{15}}{(R_{16} + \frac{R_{17}}{2})}\right) = U_{R15} + U_{R16R17} = 2.8V$$

If U_{BEK7} = 0.7V, then R_{15} must be three times the sum of R_{16} and R_{17} to get a collector-emitter voltage of 2.8V for the four output transistors. If R_{15} is three times higher also U_{R15} is three times higher than U_{R16R17}.

$$\mathbf{U_{R15}} = \frac{3}{4} * 2.8V = \mathbf{2.1V}$$

$$\mathbf{U_{R16R17}} = \frac{1}{4} * 2.8V = \mathbf{0.7V}$$

$$\mathbf{R_{15}} = \frac{U_{R15}}{I_{R15}} = \frac{2.1V}{1.4mA} = \mathbf{1.5k\Omega}$$

$$\mathbf{R_{R16R17}} = \frac{U_{R16R17}}{I_{R16R17}} \approx \frac{R_{15}}{3} = \frac{1.5k\Omega}{3} = \mathbf{500\Omega}$$

$$\mathbf{R_{16}} < 500\Omega = \mathbf{100\Omega} \rightarrow \text{assumed}$$

$$R_{16} = R_{R16R17} - \frac{R_{17}}{2} \rightarrow \mathbf{R_{17}} = 2 * (R_{R16R17} - R_{16}) = 2 * 400\Omega = 800\Omega \approx \mathbf{1k\Omega}$$

The 2.8V should be reached when the potentiometer is in its middle position (50%). That is why R_{17} is divided by two. On the market, there is no 800Ω potentiometer.

$$\mathbf{U_{CEK7}} = U_{BEK7} * \left(1 + \frac{R_{15}}{(R_{16} + \frac{R_{17}}{2})}\right) = 0.7V * \left(1 + \frac{1.5k\Omega}{600\Omega}\right) = \mathbf{2.45V}$$

With a 1kΩ potentiometer, the 2.8V are reached with 40% instead of 50%. In middle position, the U_{CEK7} would equal 2.45V.

The quiescent current is adjusted without an output load!

6 Voltage Amplification Stage (VAS)

The VAS is a class A-Amplifier with an NPN transistor. The bias collector current is delivered from the constant current source.

This VAS is not the beta-enhanced VAS according to Douglas Self's amplifiers. I did not find any formulas for a beta-enhanced VAS and according to Douglas Self, a beta-enhanced VAS can create instability of the whole amplifier, if you don't know what to do. Because of that, a normal Class A amplifier is used.

6.1 VAS emitter resistor (R_{18}):

The emitter resistor R_{18} is used as current feedback and should stabilize the VAS. The voltage drop across this resistor should be around 0.6V. The higher this voltage, the better is the current feedback. But higher U_{R18} limits the maximal output voltage.

$$\mathbf{R_{18}} = \frac{U_{R18}}{I_{CK8}} = \frac{0.6V}{9.3mA} = 64\Omega \approx \mathbf{68\Omega}$$

6.2 VAS base current:

$$\mathbf{I_{BK8}} = \frac{I_{CK8}}{\beta_{K8}} = \frac{9.3mA}{300} = \mathbf{31uA}$$

This current of 31uA is really small and is delivered from the constant current source of the differential amplifier.

6.3 Voltage Gain of the VAS (V_{VAS}):

$$\mathbf{V_{VAS}} = \frac{R_{iK6} + \mathbf{R_{R16R17}}}{R_{18}} \approx \frac{R_{iK6}}{R_{18}} = \frac{507k\Omega}{68\Omega} = \mathbf{7485}$$

The voltage gain of the VAS equals 7485 and is very high!

6.4 Miller capacitance:

C_5 reduces the gain for high frequencies (Miller capacitance) and guarantees that the amplifier is stable. Without C_5 there is a risk that the amplifier will oscillate in the MHz range and act as a jammer.

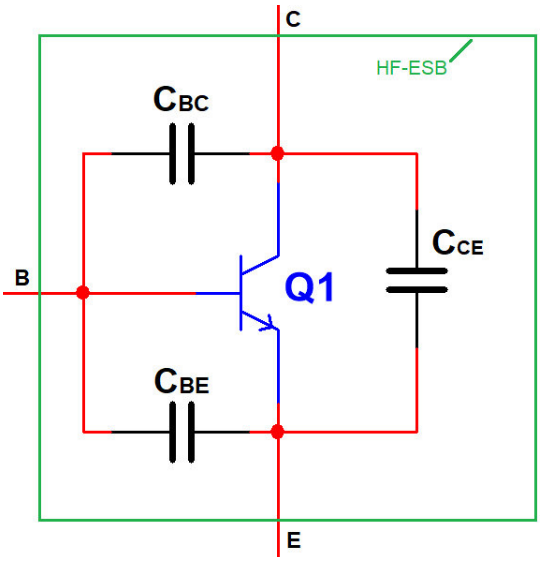
Basically, the smaller the capacitor, the higher the frequencies that are transmitted. As a result, the amplifier is more unstable. The larger the capacitor, the lower the frequencies that are transmitted. This could lead to attenuated signals in the audio band.

The upper cut-off frequency of the VAS should be around 100kHz. With a miller capacitance, an input and output cut-off frequency can be calculated. The smallest cut-off frequency determines the whole VAS cut-off frequency.

There are formulas for calculating the Miller capacitance. However, in practice, several capacity values are tried, or simulated to get the best result. 100pF is a good guideline and is also recommended by Douglas Self. Depending on the VAS gain, the 100pF may vary.

- Formulas:

To understand the formulas, some knowledge of transistor equivalent circuits and miller capacitance should be present. A detailed description is neglected.



$$C_{BCges} = C_{BC} + C_5 \approx C_5$$

$$C_{MBE} = C_{BCges} * (1 - V_{VAS}) \approx C_5 * V_{VAS}$$

$$C_{MCE} = C_{BCges} * \left(1 - \frac{1}{V_{VAS}}\right) \approx C_5$$

$$f_{coBE} = \frac{1}{2 * \pi * r_{IN} * C_{MBE}}$$

$$f_{coCE} = \frac{1}{2 * \pi * r_{OUT} * C_{MCE}}$$

7 Input Stage with a differential amplifier

7.1 Important Parameters (V_{UD}, V_{UC}, CMRR):

The differential amplifier has a really high input resistance. As a result, the amplifier does not act as a load for the preamplifier and the input filter.

It has two main specifications. The "Differential Mode Gain" (V_{UD}) and the "Common Mode Gain" (V_{UC}). The ratio of these two parameters should be high and is called "Common Mode Rejection Ratio" (CMRR).

$$\mathbf{CMRR} = \left|\frac{\mathbf{V_{UD}}}{\mathbf{V_{UC}}}\right| \rightarrow \mathbf{CMRR} \gg \rightarrow \mathbf{V_{UD}} \gg \textbf{and } \mathbf{V_{UC}} \ll$$

$$\mathbf{V_{UC}} \approx \frac{\mathbf{R_C}}{\mathbf{2 * R_E}} \rightarrow \mathbf{V_{UC}} \ll \rightarrow \frac{\mathbf{R_C}}{\mathbf{R_E}} \ll \rightarrow \mathbf{R_E} \gg$$

$$\mathbf{V_{UD}} \approx \frac{\mathbf{s * R_C}}{\mathbf{2}} = \frac{\mathbf{I_C * R_C}}{\mathbf{2 * U_T}} \rightarrow \mathbf{V_{UD}} \gg \rightarrow \mathbf{R_C} \gg \rightarrow \mathbf{I_C} >$$

In this three equation, it can be seen that a high-quality differential amplifier should have a high emitter resistor, a high collector resistor (but much smaller than R_E) and a higher collector current. Don't assume I_C too high, otherwise, the power consumption becomes unnecessarily high. The high emitter resistor is achieved with the internal resistance of a constant current source.

7.2 Constant current source for differential amplifier:

given: $+U_B = 20V$, $U_{LEDP1} = 2.1V$, $U_{BEK5} = 0.7V$, $I_{CK5} = 5mA$, $\beta_{K5} \approx 300$, $I_{LEDP1} = 10mA$;

- R_{12} – Calculation:

$$\mathbf{I_{CK5}} = \frac{U_{LEDP1} - U_{BEK5}}{R_{12}} \rightarrow R_{12} = \frac{U_{LEDP1} - U_{BEK5}}{I_{CK5}} = \frac{1.4V}{0.005A} = 280\Omega \approx 330\Omega \rightarrow I_{CK5} = \mathbf{4.24mA}$$

$$\mathbf{I_{CK1}} = \mathbf{I_{CK2}} = \frac{\mathbf{I_{CK5}}}{\mathbf{2}}$$

- R_{iK5} – Calculation (K_5):

$$\mathbf{R_{iK5}} \approx r_{CE} * \frac{\beta_{K5}}{1 + \frac{\beta_{K5} * U_T}{U_{12}}} = \frac{100V}{I_{CK5}} * \frac{\beta_{K5}}{1 + \frac{\beta_{K5} * U_T}{U_{12}}} = \frac{100V}{4.24mA} * \frac{300}{1 + \frac{300 * 25mV}{1.4V}} = \mathbf{1.1M\Omega}$$

R_{iK5} is the output resistance of the current source and should be as high as possible. The higher β, the higher is R_{iK5}.

7.3 Emitter resistors (R_6, R_7) for DC-Bias stabilization:

These two resistors should stabilize the DC operating point of K_1 and K_2. According to Douglas self, these two resistances should also drastically reduce the DC offset.

A rule of thumb for these two resistors is:

$$\mathbf{R_6} = \mathbf{R_7} = \frac{U_{R6}}{I_{CK1}} = \frac{U_{R7}}{I_{CK2}} = \frac{250mV}{\frac{I_{CK5}}{2}} = \frac{250mV}{2.12mA} \approx \mathbf{100R}$$

$$\mathbf{R_E} = R_{iK5} + R_6 \approx R_{iK5} = \mathbf{1.1M\Omega}$$

R_{iK5} is much higher than R_6 and therefore, R_6 can be neglected.

7.4 Current mirror:

- R_C – Calculation:

The DC-Bias-Voltage of 1.3V of the VAS is also at the current mirror. With this voltage and the collector current, R_C can be calculated. R_C consists of R_8 and the transistor K_3 resistance. For a current feedback, which should stabilize the current mirror the resistors R_8 and R_9 are used. The voltage drop across these resistors should be around 150mV and the current through R_8 is I_{CK3} minus I_{BK8}. I_{BK8} can be neglected because it is really small (uA).

$$\mathbf{R_C} = \frac{U_{Rc}}{I_{CK3}} = \frac{U_{Rc}}{I_{CK1}} = \frac{U_{R18} + U_{BEK8}}{I_{CK1}} = \frac{0.6V + 0.7V}{2.12mA} = \mathbf{613\Omega}$$

$$\mathbf{R_8} = \mathbf{R_9} = \frac{U_{R8}}{I_{CK3} - I_{BK8}} \approx \frac{U_{R8}}{I_{CK3}} = \frac{U_{R9}}{I_{CK4}} = \frac{150mV}{2.12mA} = 70.75\Omega \approx \mathbf{68\Omega}$$

- Mirrored current:

$$\mathbf{I_{CK3}} = I_{CK4} * \frac{R_9}{R_8} = I_{CK4} * \frac{68\Omega}{68\Omega} = I_{CK4} = \mathbf{2.12mA}$$

7.5 Calculate V_{UD}, V_{UC}, CMRR:

$$\mathbf{V_{UC}} \approx \frac{R_C}{2 * R_E} = \frac{613\Omega}{2 * 1.1M\Omega} = \mathbf{0.000279}$$

$$\mathbf{V_{UD}} \approx \frac{s * R_C}{2} = \frac{I_C * R_C}{2 * U_T} = \frac{2.12mA * 613\Omega}{2 * 25mV} = \mathbf{26}$$

$$\mathbf{CMRR} = \left|\frac{V_{UD}}{V_{UC}}\right| = \left|\frac{26}{0.000279}\right| = 93280 = \mathbf{99.4db}$$

A CMRR of 99.4dB should be enough for a good quality amplifier input.

7.6 Input resistance of the differential amplifier:

$$\mathbf{r_{IN}} = 2 * (\beta + 1) * (r_{eK1} + R_E) \approx 2 * \beta * \left(\frac{U_T}{I_{CK1}} + R_E\right) = 2 * 300 * \left(\frac{25mV}{2.12mA} + 1.1M\Omega\right) \approx \mathbf{660M\Omega}$$

As you can see, the input impedance of transistor differential amplifier depends on the emitter resistor and the current gain. Because of the high input resistance of the current source and a current gain of β ≈ 300, an input impedance of 660MΩ occurs.

7.7 Input RC-Filter:

The input filter is an RC high pass filter and should filter remaining DC-Offset from the preamplifier. The input resistor R_{19} should be in a range of 1kΩ to 10kΩ and should reduce the noise sensitivity at the input. If it is too high (e.g. 100kΩ), the amplifier is more noise sensitive. R_{19} has no impact to the input signal frequency, only C_1 and R_5 set the high pass cut-off frequency f_{cl}. Also, a low pass filter can be installed at the amplifier input. However, the input signal of this amplifier is filtered with a 2nd order active filter network in the pre-amplifier. The high input resistance of the differential amplifier has no impact on the RC-Filter and can be neglected.

given: C_1 = 10uF, R_5 = 10kΩ;

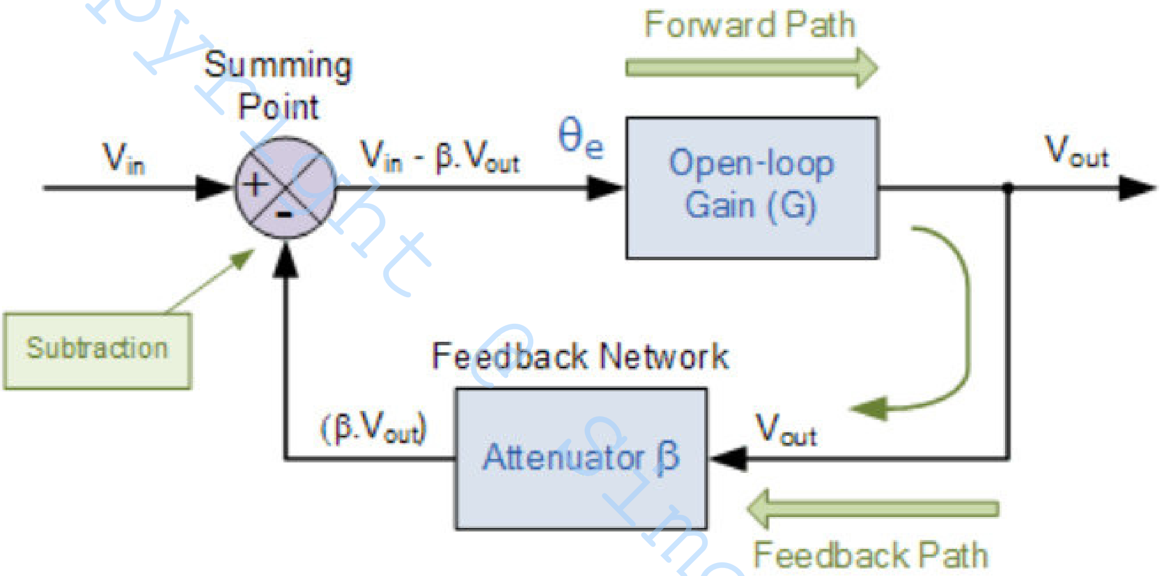
$$\mathbf{f_{cl}} = \frac{1}{2 * \pi * C_1 * R_5} = \mathbf{1.59Hz}$$

Keep in mind, that an electrolytic capacitor doesn't withstand negative voltage. Therefore, C_5 should be set on the half supply voltage. In this circuit R_4 and R_5 equals 10kΩ and set C_1 on the half supply voltage. In the chapter DC-Offset correction a resistor network (R_1, R_2 and R_3) is added, to reduce the DC-Offset at the amplifier output. In the end, a 10uF ceramic SMD capacitor was chosen for C_5.

8 Negative feedback (NFB)

8.1 Theory of a control loop:

The open-loop gain (G) should be high to push the signals as fast as possible through the amplifier to improve the slew rate. At the output, a fraction of the signal is fed backwards in the differential amplifier to decrease the overall amplification (closed loop gain G_C). Now, the slew rate is high enough and the output signal is not too high and is not distorted by the output transistors. The NFB works like a control circuit and stabilizes the whole amplifier.



8.2 Open-loop gain (G):

$$\mathbf{G} = V_{UD} * V_{VAS} * V_{UDS} * V_{UOS} = 26 * 7485 * 1 * 1 = \mathbf{194617} = 105\text{dB}$$

8.3 Closed-loop gain (G_C):

$$\mathbf{G_C} = \frac{\mathbf{V_{OUT}}}{\mathbf{V_{IN}}} = \frac{\mathbf{u_{OUT}}}{\mathbf{u_{IN}}} = \frac{\mathbf{G}}{\mathbf{1 + \beta * G}}$$

V_{OUT}, V_{IN} and G are known, β has to be calculated. The factor β is set with a resistor divider (R_{10} and R_{11}). To make β frequency-dependent, two capacitors (C_3 and C_4) are added. This four components act as a bandpass filter. At frequency outside 20Hz to 20kHz, β is rising and the closed loop gain is decreasing. With this method, oscillation in unwanted frequency ranges can be avoided.

8.4 Calculate (R_{10} and R_{11}):

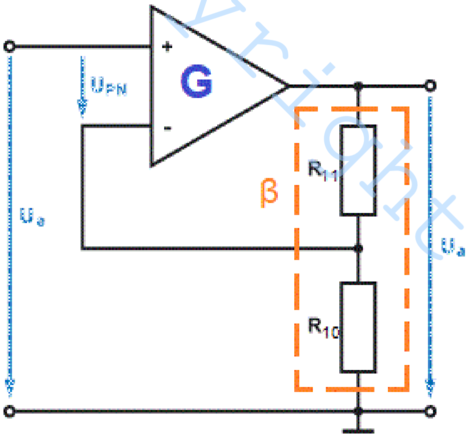
Given.: $u_{OP} = 19.8V_p$, $u_{IN} = 2.32V_{RMS}$, $G = 194617$;

$$\mathbf{u_{OUT}} = \frac{u_{OP}}{\sqrt{2}} = \frac{19.8V}{\sqrt{2}} = \mathbf{14V_{RMS}}$$

$$G_C = \frac{u_{OUT}}{u_{IN}} = \frac{G}{1 + \beta * G} \rightarrow \boldsymbol{\beta} = \frac{\mathbf{U_{IN} * G - U_{OUT}}}{\mathbf{U_{OUT} * G}}$$

$$\boldsymbol{\beta} = \frac{u_{IN} * G - u_{OUT}}{u_{OUT} * G} = \frac{2.32V * 194617 - 14V}{14V * 194617} = \mathbf{0.166}$$

Compared to a non-inverting Op-amp circuit, R_{10} and R_{11} can be calculated like:



$$\boldsymbol{\beta} = \frac{\mathbf{R_{10}}}{\mathbf{R_{10} + R_{11}}}$$

R_{10} is assumed with 10kΩ.

$$\mathbf{R_{11}} = \frac{R_{10}}{\beta} - R_{10} = \frac{10k\Omega}{0.166} - 10k\Omega = 50k\Omega \approx \mathbf{47k\Omega}$$

$$\boldsymbol{\beta} = \frac{R_{10}}{R_{10} + R_{11}} = \frac{10k\Omega}{57k\Omega} = \mathbf{0.175}$$

$$\mathbf{u_{OUT}} = \frac{u_{IN} * G}{1 + \beta * G} = \frac{2.3214V_{RMS} * 194617}{1 + 0.175 * 194617} = 13.26V_{RMS} = \mathbf{18.75V_p}$$

If the open-loop gain is very high, a simplification for the closed-loop gain can be derived.

$$G_C = \frac{G}{1 + \beta * G}$$

$$\beta = \frac{1}{G_C} - \frac{1}{G} \approx \frac{1}{G_C}$$

$$\mathbf{G_C} = \frac{1}{\beta} = \frac{1}{\frac{R_{10}}{R_{10} + R_{11}}} = \frac{R_{10} + R_{11}}{R_{10}} = 1 + \frac{R_{11}}{R_{10}} = 1 + \frac{47k\Omega}{10k\Omega} = \mathbf{5.7}$$

$$\frac{\mathbf{u_{OUT}}}{\mathbf{u_{IN}}} = \mathbf{G_C} = \mathbf{1} + \frac{\mathbf{R_{11}}}{\mathbf{R_{10}}}$$

This formula is well-known in non-inverting Op-amp circuits.

8.5 Calculate bandpass filter (C_3 and C_4):

R_{10} and C_3 specify the lower cut-off frequency (f_{cl}) and should be lower 20Hz. R_{11} and C_4 specify the upper cut-off frequency (f_{co}) and should be around 100kHz.

- Low pass filter:

$$\mathbf{f_{cl}} = \frac{1}{2 * \pi * R_{10} * C_3} = \frac{1}{2 * \pi * 10k\Omega * 100uF} = 0.\mathbf{159Hz}$$

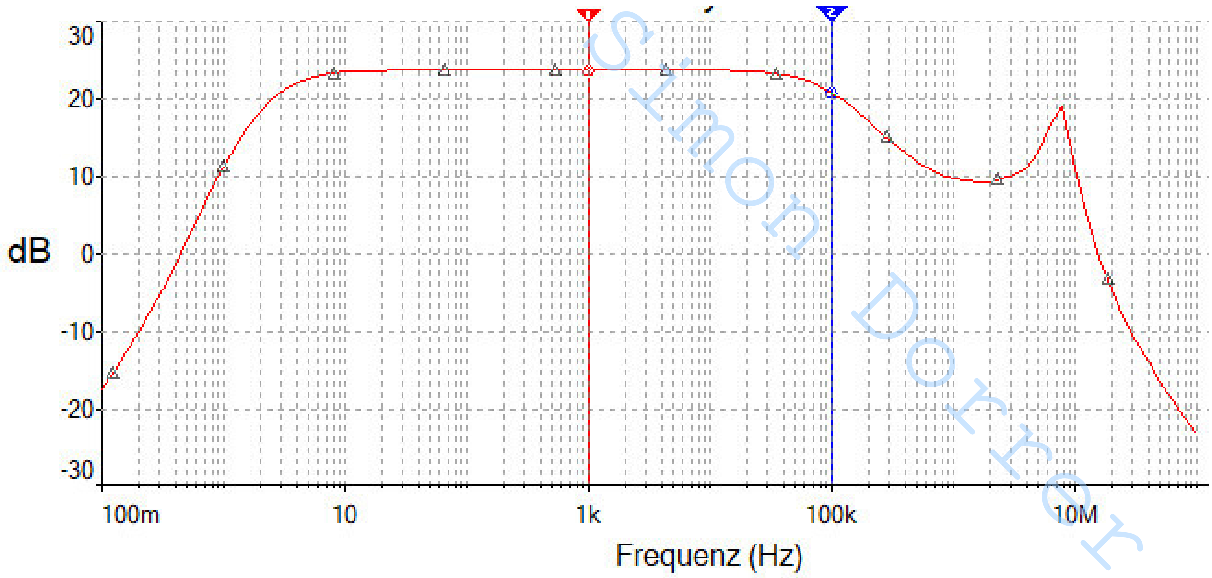
For the low pass filter, 100uF for C_3 is assumed. With that capacitance, a lower cut-off frequency of 0.159Hz is achieved.

- High pass filter:

$$\mathbf{C_4} = \frac{1}{2 * \pi * R_{11} * \mathbf{f_{co}}} = \frac{1}{2 * \pi * 47k\Omega * 100kHz} = \mathbf{33pF}$$

For the high pass filter, 100kHz for f_{co} is assumed. This frequency is achieved with a 33pF capacitor for C_4. **Attention! In some cases, the amplifier oscillates with C_4. Try other values or leave it away!**

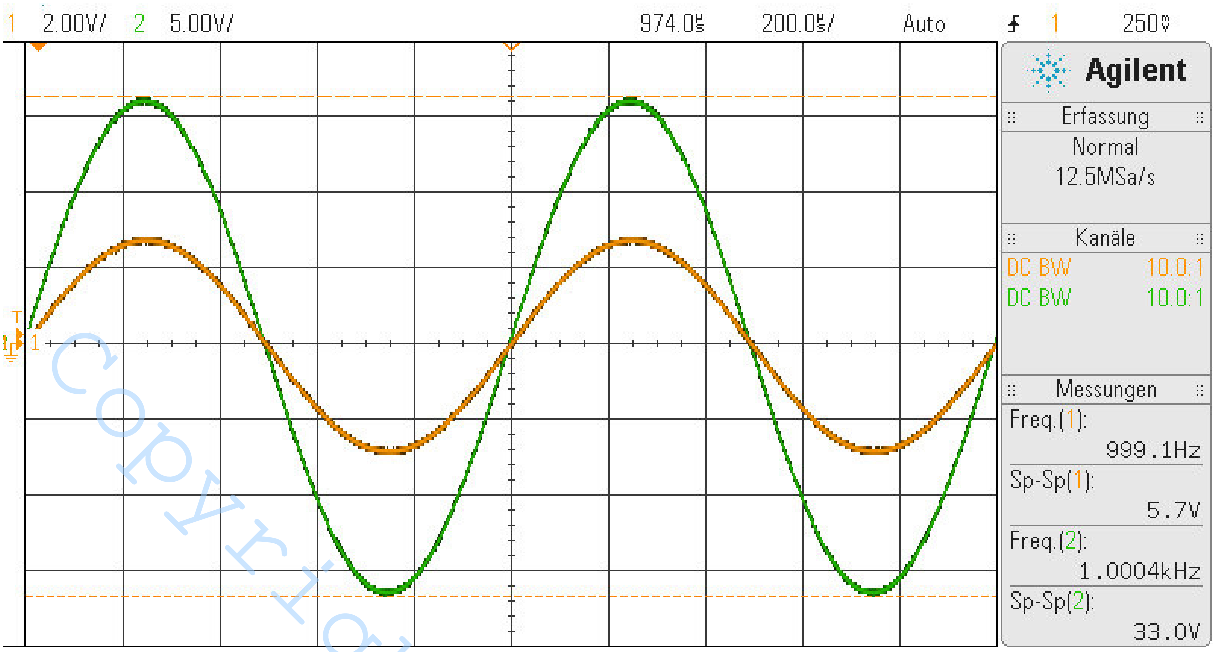
8.6 Transfer function with negative feedback:



It can be seen, that the transfer function has a lower cut-off frequency of 3Hz and an upper cut-off frequency of 101.4kHz. This bode diagram verifies the functionality of the negative feedback.

Furthermore, at around 8MHz, the resonance frequency of the speaker can be seen. This spike may be fed back into the amplifier and may cause oscillation. How to get rid of this, is explained under Output filter network.

8.7 Measurement of input and output peak voltage:



$$G_C = \frac{u_{OUT}}{u_{IN}} = \frac{33Vpp}{5.7Vpp} = \mathbf{5.79} \quad \rightarrow \quad G_C = 1 + \frac{R_{11}}{R_{10}} = 1 + \frac{47k\Omega}{10k\Omega} = \mathbf{5.7}$$

With this oscilloscope picture, the calculation of the closed loop gain in 8.3 can be confirmed. 33Vpp is the biggest peak-peak output voltage in a 4Ω load.

9 Oscillating amplifiers

It often happens that an amplifier outputs a periodic signal without an input signal. In this case, the amplifier has become an unwanted oscillator. An oscillator is an electronic circuit that produces a periodic, oscillating electronic signal, often a sine wave or a square wave.

There are two conditions, which must be not met, otherwise, the amplifier will become an oscillator. **Only when both conditions are met, the amplifier oscillates.**

9.1 Amplitude condition:

$$\mathbf{G * \beta \approx 1 \dots 3}$$

If the multiplication factor of the open-loop gain (G) and the feedback gain (β) equals 1-3, the oscillator amplitude condition is met.

9.2 Phase shift condition:

$$\boldsymbol{\varphi_G + \varphi_\beta = 0° \text{ or } 360°}$$

If the sum of the open-loop phase shift (φ_G) and the feedback phase shift (β) equals 0°, the amplifier may oscillator.

In most amplifier applications, the phase shift in the used frequency range is 0°. Therefore, care must be taken that the amplitude condition is not met and is lower than 1, otherwise, the amplifier oscillates.

10 Output filter network

10.1 Purpose of the RL-Network:

The RL network is supposed to let the audio signal through, but will not let resonance frequencies from the speaker go back into the power amp. So, this network acts as a decoupling network for high frequencies.

Douglas Self has tried many RL variants and has come to a very good result with 2.3uH and 10Ω. With the resistor, the Q can be adjusted and with the coil, the frequency can be shifted.

10.2 Create Inductor:

According to Douglas Self, an inductivity of 2.3uH is adequate for most designs. This is achieved with a 1mm thick wire, winded 16 times around a 15mm thick AA battery. Keep attention to the maximum current rating of your wire. For instance, if you build a monster power amp, 1mm thick wire may not be thick enough. Your coil can be designed as you wish. I have calculated these values on the following website:

https://www.electronicdeveloper.de/InduktivitaetLuftEinl.aspx

10.3 Power loss of R_{24}:

The impedance of the coil depends on the frequency. For the worst case, f is 20kHz.

$$\mathbf{R_{L1}} = X_{L1} + R_{L1DC} = 2 * \pi * f * L_1 + R_{L1DC} = 2 * \pi * 20kHz * 2.3uH + 18m\Omega = \mathbf{0.29\Omega}$$

$$\mathbf{i_{R24}} = i_{CK12P} * \frac{R_{L1}}{R_{24} + R_{L1}} = 4.69A * \frac{0.29\Omega}{10\Omega + 0.29\Omega} = \mathbf{0.132mA}$$

$$\mathbf{P_{24max}} = i_{R24}^2 * R_{24} = (0.132mA)^2 * 10\Omega = \mathbf{0.174W}$$

It can be seen, that the current through R_{24} is very small and a 0.25W resistor can be taken.

10.4 Purpose of the RC network (Zobel network):

A capacitor weakens the gain of the NFB in a certain range, so a coil amplifies it. At high NFB gain, the amplifier can become unstable. Therefore, the Zobel network serves as a "reactive current compensation". The positive reactance, which is contained in the speaker impedance (= inductive component), is compensated by the negative reactance of the RC element. This generates a real ohmic resistance at the amplifier output.

10.5 Calculate Zobel network:

In general, for the Zobel network, there is no 100% correct calculation available. However, there are good approximations:

given.: Visaton FR-10: R_{LS} = 3.3Ω, L_{LS} = 0.2mH;

$$\mathbf{R_{23}} = 1.25 * (R_{LS} + R_{24}) \approx 1.25 * R_{LS} = 1.25 * 3.3\Omega = 4.13\Omega \approx \mathbf{4.3\Omega}$$

$$\mathbf{C_9} = \frac{L_{LS} + L_1}{(R_{23})^2} \approx \frac{L_{LS}}{(R_{23})^2} = \frac{0.2mH}{(4.3\Omega)^2} = 13.8uF \approx \mathbf{10uF}$$

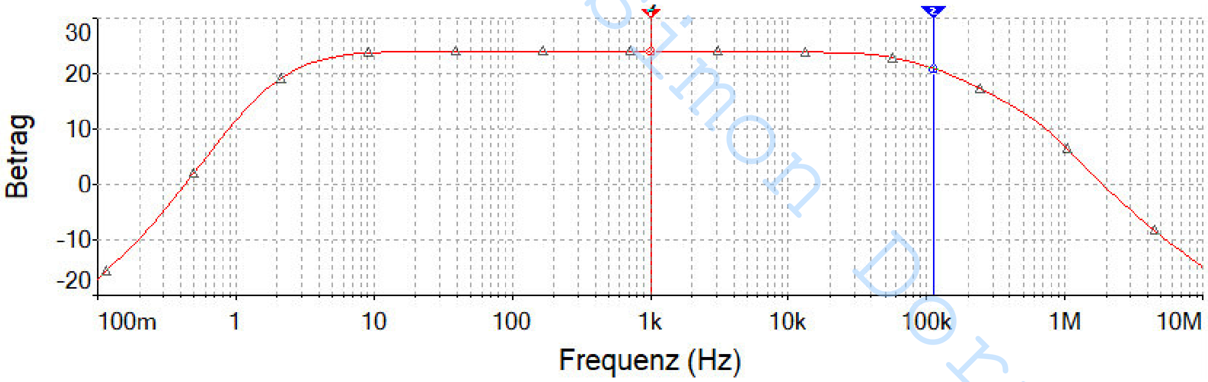
The derivation of the C_9 formula can be found in the attached .pdf file.

In most cases, the speaker inductance is much higher than the 2.3uH of the decoupling coil and can be neglected.

As described above, R_{24} has nearly no influence on the load at low frequencies, because it is shorted by the coil and can be neglected.

10uF unipolar capacitors aren't that popular. So, if your calculated components are not available, a 10Ω and a 100nF capacitor can be taken as well.

10.6 Transfer function with output filter network:



Compared to the transfer function under Negative feedback, this transfer function is not disturbed by the speaker. Therefore, the functionality of the coil and the Zobel network can be confirmed.

11 DC-Offset correction

DC-Offset at the output can destroy the speaker and make sound quality worse. Therefore, a resistor network (R_1, R_2 and R_3) with one potentiometer and two resistors between $+U_B$ und $-U_B$ is used.

The voltage across the potentiometer (R_2) should be between +-1V. With this voltage, a high enough DC-Offset range can be corrected. If you choose a higher voltage, the potentiometer value gets higher and the adjustment is getting more difficult. So, R_1 and R_3 have to be designed carefully to fulfill this requirement. The current through the resistor network shouldn't be too high and is assumed with 0.5mA.

given.: U_{R2} = +-1V = 2V, I_{R123} = 0.5mA;

$$\mathbf{R_2} = \frac{U_{R2}}{I_{R123}} = \frac{2V}{0.5mA} = 4000\Omega \approx \mathbf{3.9k\Omega}$$

When the potentiometer is in its middle position, the 2V should be split exactly in half (1V).

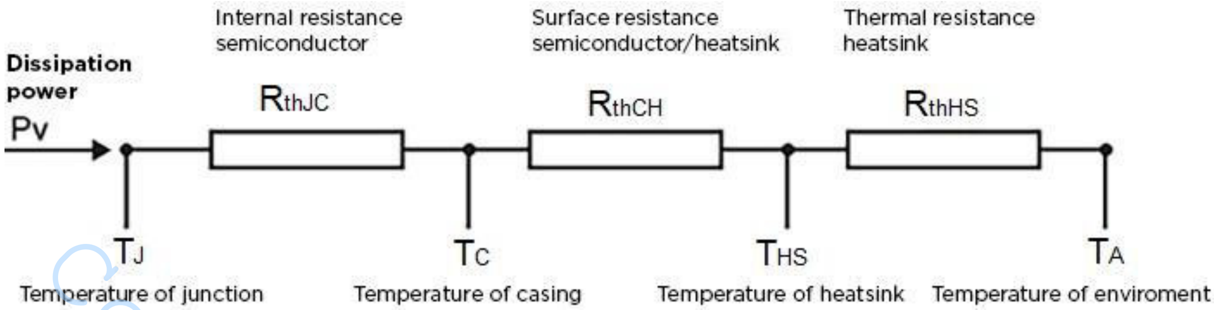
Thus, 20V-1V = 19V should drop across the resistors R_1 and R_3.

$$\mathbf{R_1} = \mathbf{R_3} = \frac{U_{R1}}{I_{R123}} = \frac{U_{R3}}{I_{R123}} = \frac{19V}{0.5mA} = 38k\Omega \approx \mathbf{39k\Omega}$$

The DC-offset is adjusted with an output load!

12 Heatsink calculation

12.1 Thermal equivalent circuit



This circuit shows the thermal resistors and their corresponding temperatures. With this circuit, every thermal formula can be derived.

12.2 Calculate max. power loss without heatsink:

$$P_{Vmax} = \frac{T_J - T_A}{R_{thJA}} = \frac{150°C - 45°C}{35.7\frac{K}{W}} = \mathbf{2.941W < 9.48W}$$

The maximum power loss of one transistor equals 9.48W and is higher than the maximum power dissipation without a heatsink → **Heatsink needed!**

12.3 Calculate thermal resistance of heatsink:

R_{thJC} is taken from the datasheet of the 2SC5200. R_{thCH} is assumed with 1.5K/W for mica discs (glimmer).

$$\mathbf{R_{thHS}} = \frac{\mathbf{T_J - T_A}}{\mathbf{P_{K12max}}} - (\mathbf{R_{thJC} + R_{thCH}})$$

$$\mathbf{R_{thHS}} = \frac{150°C - 45°C}{9.48W} - \left(0.83\frac{K}{W} + 1.5\frac{K}{W}\right) < \mathbf{8.75\frac{K}{W}}$$

The heat sink of one transistor must have a thermal resistance of smaller 8.75 K/W.

12.4 Calculate temperature of heatsink:

$$\mathbf{T_{HSmax} = T_J - P_{K12max} * R_{thJC} - P_{K12max} * R_{thCH}}$$

$$\mathbf{T_{HSmax}} = 150°C - 9.48W * 0.83\frac{K}{W} - 9.48W * 1.5\frac{K}{W} = \mathbf{127.9°C}$$

The heat sink of one transistor can reach a temperature up to 127.9°C.

12.5 Several identical transistors on one heatsink:

For example, in an AB output stage, two transistors must be placed on one heatsink. If you want to build a stereo power amplifier with only one heat sink, four transistors are placed on one heat sink.

The total thermal resistance of the heatsink is a parallel circuit of all individual thermal resistances of the heatsink:

$$\frac{1}{R_{thHSges}} = \frac{1}{R_{thHS1}} + \frac{1}{R_{thHS2}} + \cdots + \frac{1}{R_{thHSn}}$$

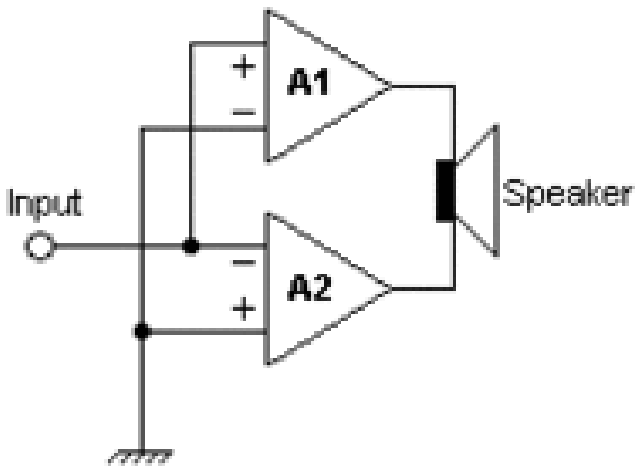
$$\frac{1}{R_{thHSges}} = 4 * \frac{1}{9.48} \rightarrow R_{thHSges} < 2.37\frac{K}{W}$$

With four transistors on one heatsink, a $R_{thHSges}$ of smaller 2.37K/W is calculated. This is a very low value and a big heatsink is needed.

To simplify your heatsink calculation, use online calculators like this one: http://www.elektronik-bastler.info/stn/kuehl.html

13 Bridged amplifiers

With this technology, you can convert a stereo amplifier into a mono amplifier with four times the normal output power. The input signal has to be inverted for the second amplifier. Both amplifiers see 2Ω instead of 4Ω load and outputs twice of the normal power. Together, they can output four times the normal output power. Keep in mind, that the amplifier can drive a 2Ω load.



14 Slew Rate (SR)

The slew rate indicates the minimum slope (V/s) of a square wave signal with the amplitude (A) at a certain frequency (f). The larger the slew rate, the higher the signal frequencies that can be transmitted without distortion.

14.1 Calculation of minimal SR:

SR… Slew Rate in V/us, A… Amplitude of the output signal

given.: f = 20kHz, A = 16.5Vp;

$$U_A = A * \sin(\omega * t) \ /\text{mit } \omega = 2 * \pi * f_{max}$$

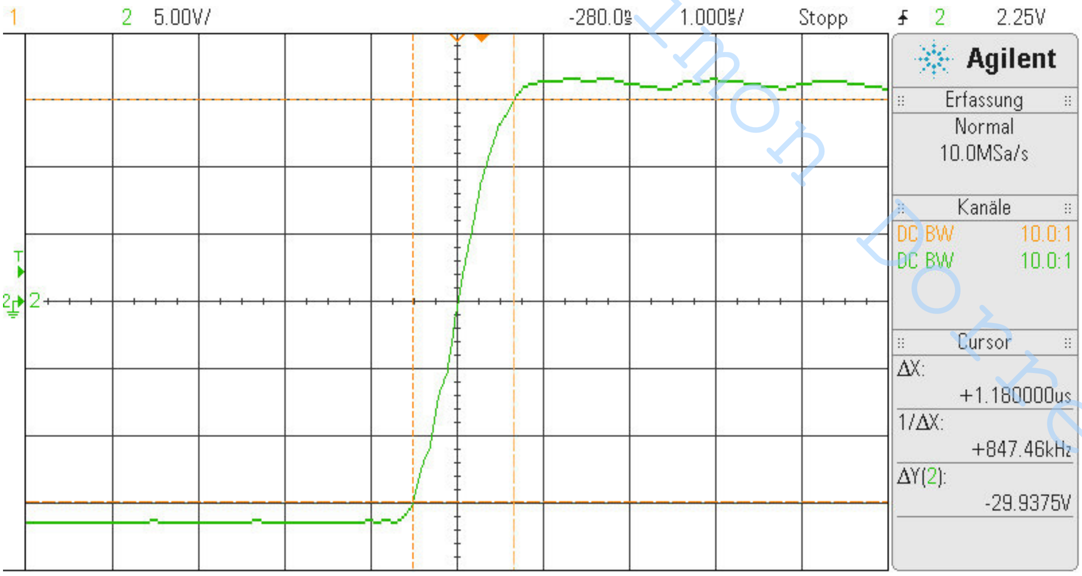
$$U_A' = \frac{dU_A}{dt} = A * \cos(2 * \pi * f_{max} * t) * 2 * \pi * f_{max} = SR$$

$$SR = A * \cos(2 * \pi * f_{max} * t) * 2 * \pi * f_{max} / \text{ mit } t = 0$$

$$SR \geq 2 * \pi * f_{max} * A = 2 * \pi * 20kHz * 16.5V \rightarrow \mathbf{SR \geq 2\frac{V}{us}}$$

The slew rate has to be higher than 2V/us to amplify sine waves with a frequency of 20kHz and an amplitude of 16.5Vp without any slew rate distortion.

14.2 Measuring SR:



$$\mathbf{SR} = \frac{dV}{dt} = \frac{30V}{1.18us} = \mathbf{25.4\frac{V}{us}}$$

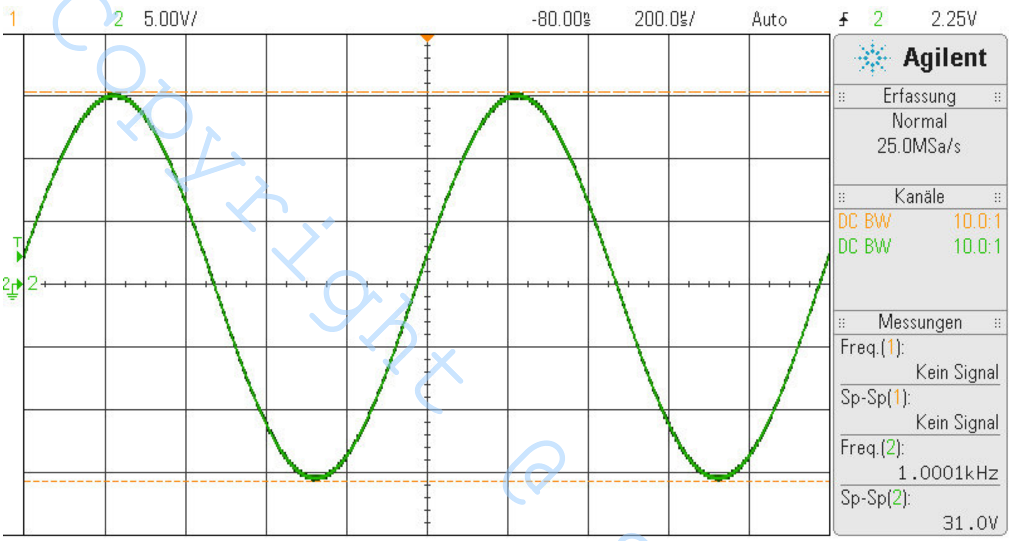
It can be seen, that the slew rate is 25.4 V/us and is high enough to amplify sine waves with a frequency of 20kHz without any distortion.

14.3 Calculation of maximal frequency:

$$\mathbf{f_{max}} = \frac{SR}{2 * \pi * A} = \frac{\frac{25.4V}{us}}{2 * \pi * 16.5V} \approx \mathbf{245kHz}$$

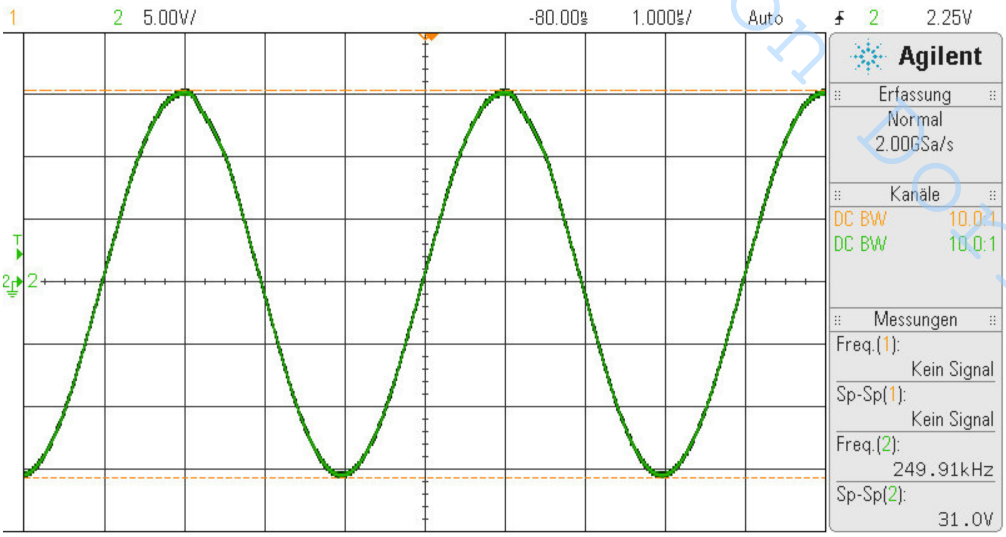
An output signal with 16.5Vp and a maximal frequency of 245kHz can be amplified without any distortion.

14.4 Output signal without slew rate distortion:



At 1kHz, no slew rate distortion can be seen.

14.5 Output signal with slew rate distortion:



At 250kHz, the positive peaks are slightly distorted. The higher the frequency, the more the sine wave will distort. However, 250kHz are not relevant for the audio frequency domain.

15 Measurements

15.1 Power recalculation:

The power calculations with the explanations under Class AB output stage are done with assumed values. Here the power is calculated with the final measured values.

15.1.1 Peak and RMS output power:

The peak output power (P_{OP}) equals:

$$\mathbf{P_{OP}} = \frac{1}{2} * \frac{u_{OP}^2}{(R_L + R_{21})} = \frac{1}{2} * \frac{(16.5V)^2}{(4\Omega + 0.22\Omega)} = \mathbf{33W_{PEAK}}$$

The RMS output power (P_{ORMS}) equals:

$$\mathbf{P_{ORMS}} = \frac{P_{OP}}{\sqrt{2}} = \mathbf{22.81W_{RMS}}$$

15.1.2 Supply power / transformer power:

$$\mathbf{P_S} = \frac{2 * +U_B * u_{OP}}{\pi * (R_{21} + R_L)} = \frac{2 * 20V * 16.5V}{3.14 * 4.22\Omega} = \mathbf{49.78W}$$

15.1.3 Peak power loss of the emitter resistor:

The peak resistor power loss (P_{R21P}) equals:

$$\mathbf{P_{R21P}} = \mathbf{P_{R22P}} = P_{OP} * \frac{R_{21}}{(R_{21} + R_L)} = 33W_{PEAK} * \frac{0.22\Omega}{4.22\Omega} = \mathbf{1.72W_{PEAK}}$$

The RMS resistor power loss (P_{R21RMS}) equals:

$$\mathbf{P_{R21RMS}} = \mathbf{P_{R22RMS}} = \frac{P_{R21P}}{\sqrt{2}} = \mathbf{1.22W_{RMS}}$$

15.1.4 Power loss of the transistor:

The peak transistor power loss (P_{K12P}) equals:

$$\mathbf{P_{K12P}} = \mathbf{P_{K10P}} = \frac{1}{2} * (P_s - P_{OP} - P_{R21P}) = \frac{1}{2} * (49.78W - 33W - 1.72W) = \mathbf{7.5W}$$

The RMS transistor power loss (P_{K12RMS}) equals:

$$\mathbf{P_{K12RMS}} = \mathbf{P_{K10RMS}} = \frac{P_{K12P}}{\sqrt{2}} = \mathbf{5.3W}$$

15.1.5 Power efficiency:

$$\boldsymbol{\eta} = \frac{P_{OP}}{P_S} = \frac{33W}{49.78W} = 0.663 = \mathbf{66.3\%}$$

Inserting the formulas for P_{OP} and P_S another formula can be found:

$$\boldsymbol{\eta} = \frac{P_{OP}}{P_S} = \frac{\pi}{4} * \frac{u_{OP}}{+U_B} = \frac{\pi}{4} * \frac{16.5V}{20V} = 0.663 = \mathbf{66.3\%}$$

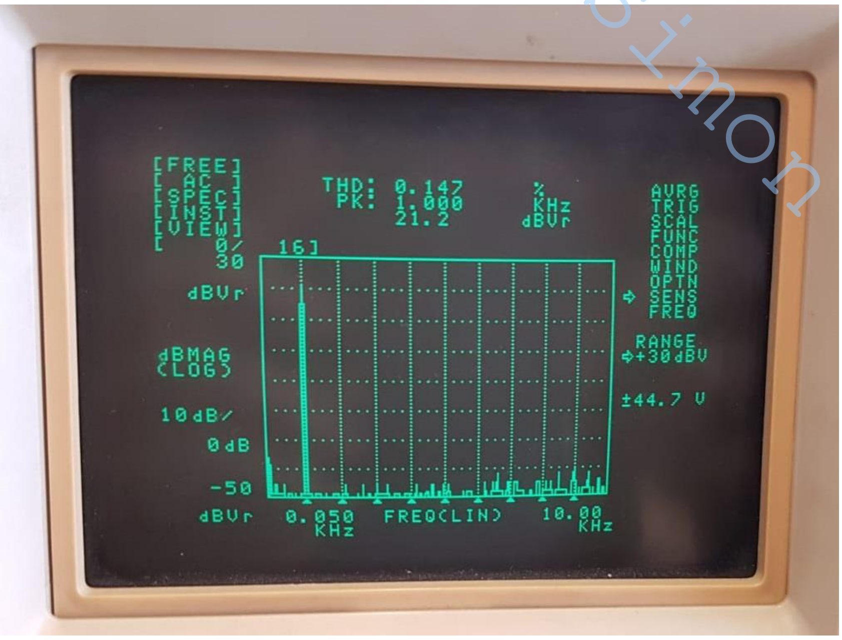
This formula is load-independent, but u_{op} has to be used with load. Because without the load, u_{op} is almost 20V and η would be wrong.

The final power efficiency equals 66.3%, which is really good for an AB-Amplifier.

15.2 Total harmonic distortion (THD):

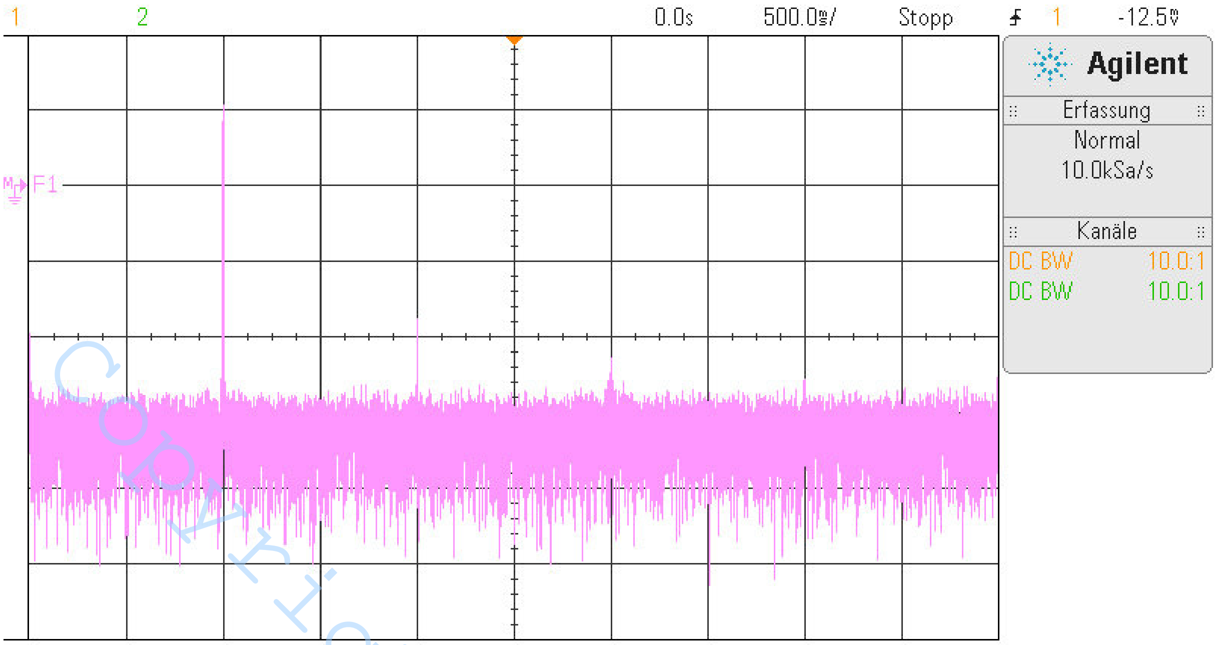
The THD was measured with the biggest output voltage (33Vpp) in a 4Ohm load. So the following results give the worst THD. With smaller output voltages, the THD is also smaller.

15.2.1 THD measured with analogue FFT analyser:



The analogue FFT analyser measured a THD of 0.147%. This THD is really good for an AB-Amplifier.

15.2.2 THD measured with oscilloscope:



$c_1 = 21.250\text{dBV} \rightarrow c_1 = 11.55\text{Veff}$ … Grundwelle @ 1kHz

$c_2 = -35\text{dBV} \rightarrow c_2 = 0.018\text{Veff}$ … 1. Oberwelle @ 2kHz

$c_3 = -45.625\text{dBV} \rightarrow c_3 = 0.0052\text{Veff}$ … 2. Oberwelle @ 3kHz

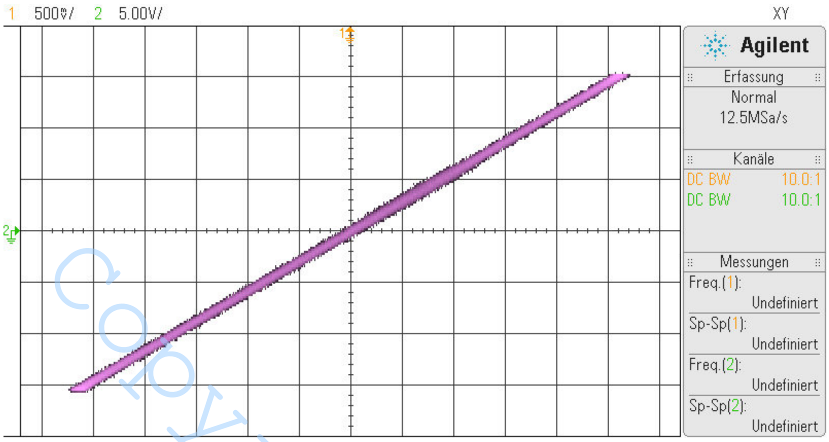
$$\mathbf{k} = \frac{\sqrt{c_2^2 + c_3^2 + \cdots + c_n^2}}{\sqrt{c_1^2 + c_2^2 + c_3^2 + \cdots + c_n^2}} = \frac{\sqrt{(0.018\text{Veff})^2 + (0.0052\text{Veff})^2}}{\sqrt{(11.55\text{Veff})^2 + (0.018\text{Veff})^2 + (0.0052\text{Veff})^2}}$$

$$\approx \mathbf{0.16\%}$$

This calculation confirms the above-measured THD with the analogue FFT analyser. Specialists can hear a THD of 0.1% of one single tone. Thus, nobody would hear a THD of 0.147%/0.16% with multiple different tones at the same time.

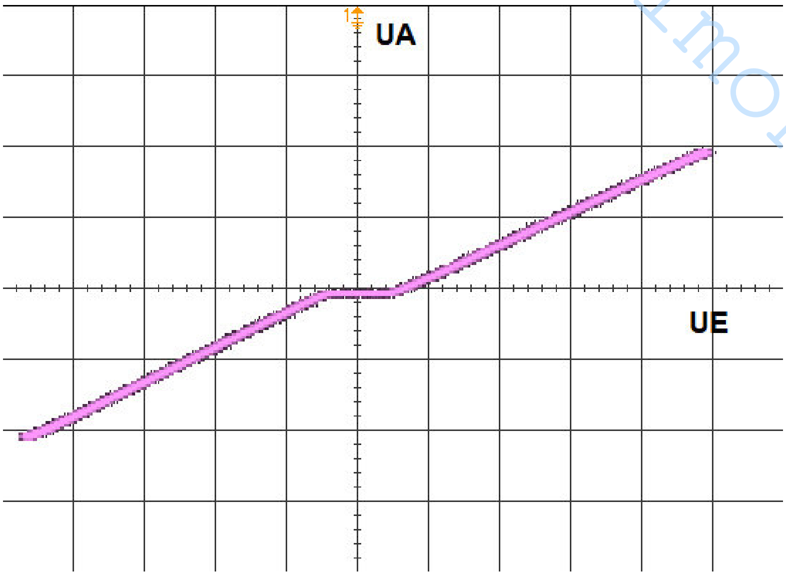
15.3 Class AB-Amplifier Crossover Distortion:

15.3.1 Adjusted AB-Amplifier:



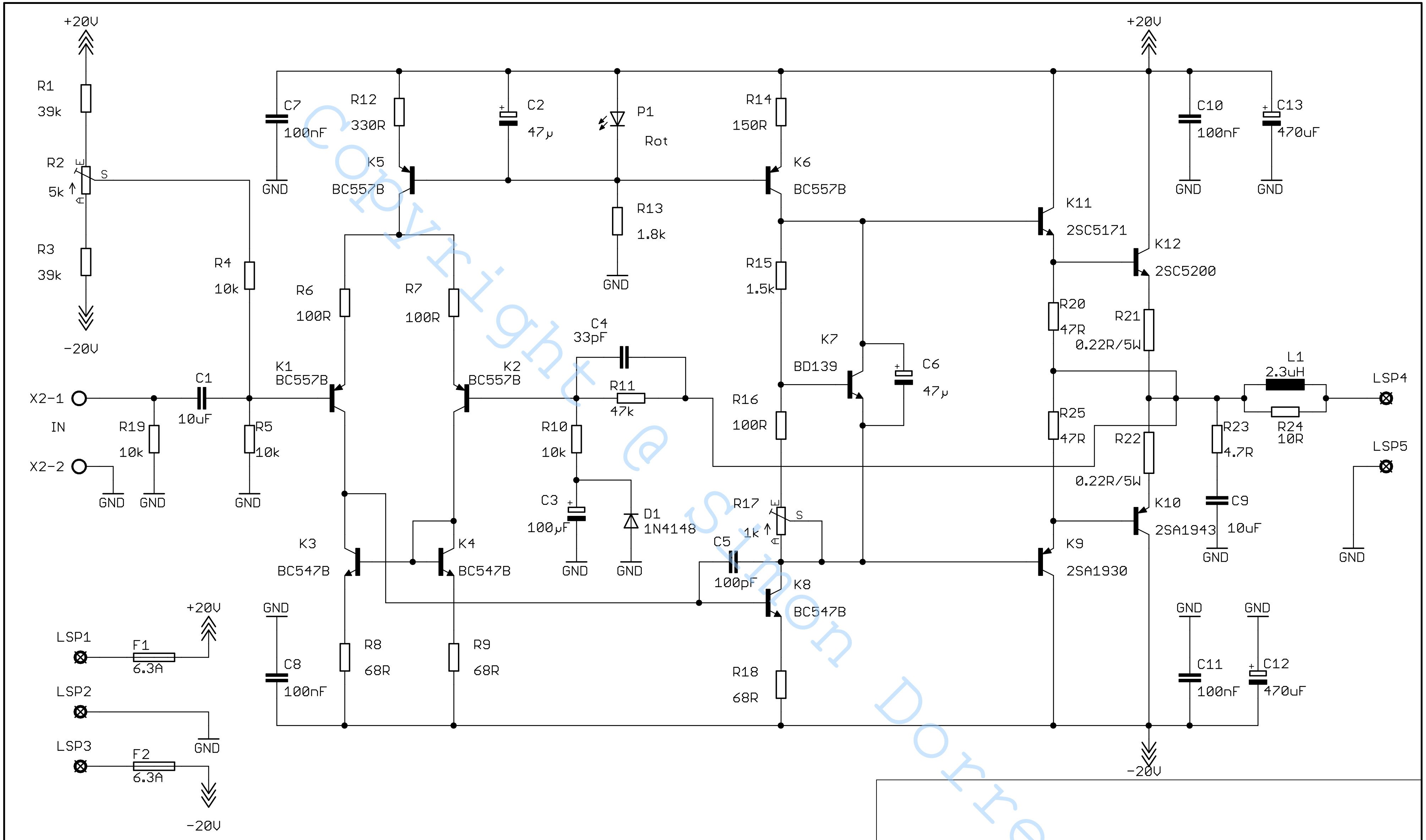
This oscilloscope pictures shows the adjusted AB-Amplifier crossover distortion in x-y-mode. On the x-axis the input voltage is applied. On the y-axis the output is shown. An ideal amplifier has a 100% linear transfer characteristic. This amplifier works almost linear without any square and cubic proportions.

15.3.2 Not adjusted AB-Amplifier (B-Amplifier):



This oscilloscope picture shows the crossover distortion of a not adjusted AB-Amplifier or a B-Amplifier. Within +-0.6V of the input signal, the output signal is 0V.

16 Circuit, PCB, BOM and 3D files



TITLE: 30W-AB-Amplifier		
Document Number:		REV:
Date: 26/01/2019 16:17:34	Sheet: 1/1	

Parts List for

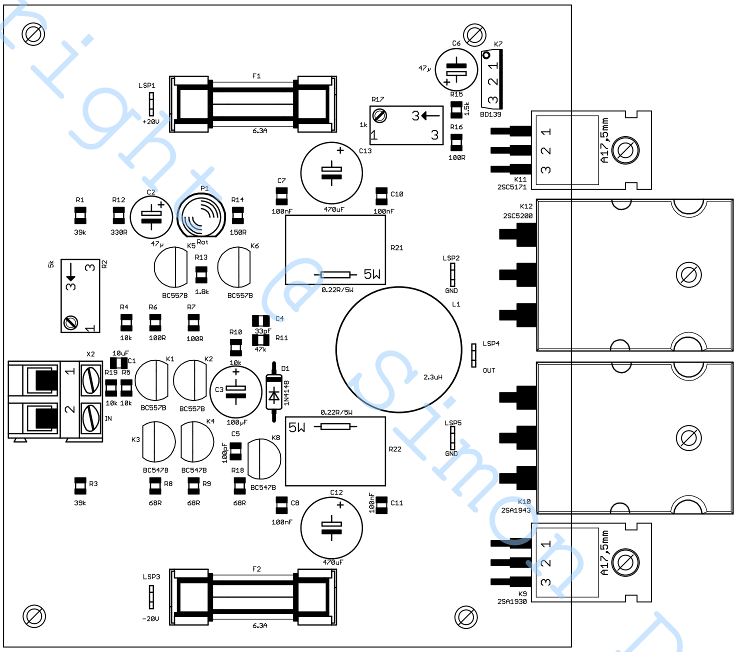
E:/HTL-Leonding/5AHEL/LA/Laborarbeit/30W-AB-Amplifier/30W-AB-Amplifier.brd

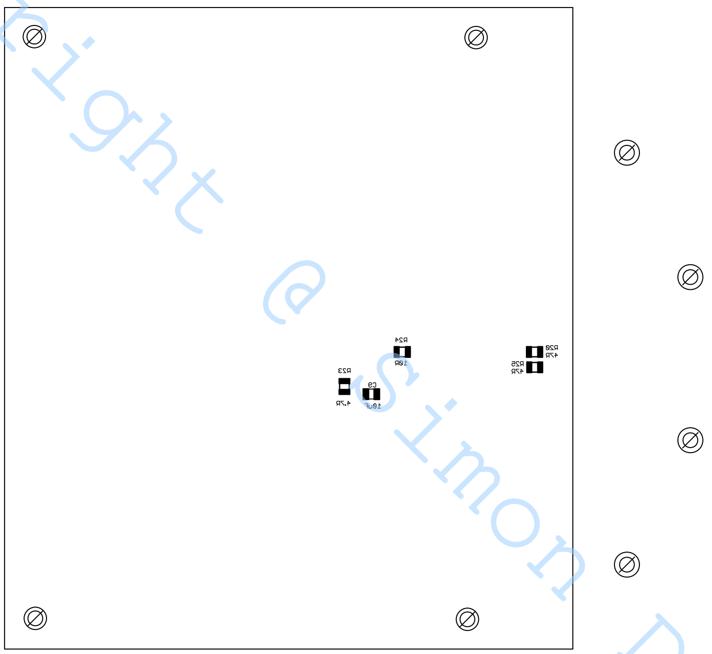
Parts Listing

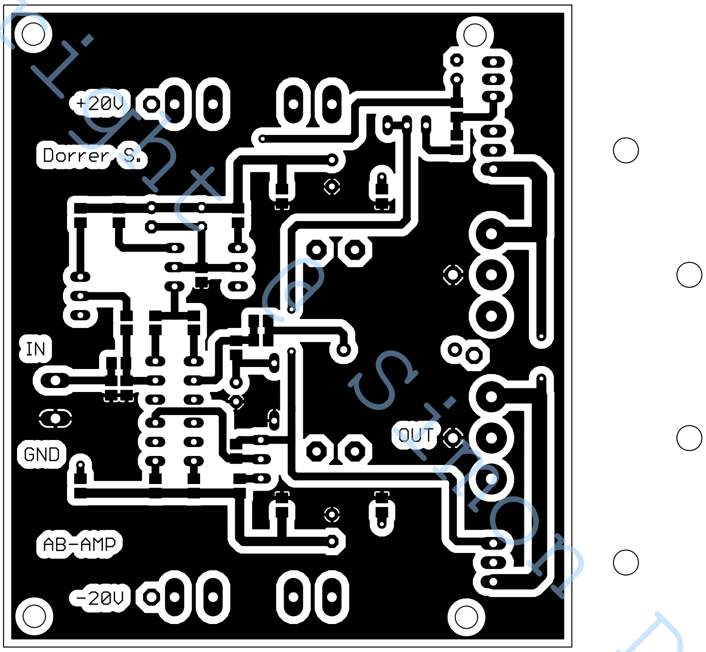
Part	Value	Package	Library
C1	10uF	C0805	rcl
C2	47µ	E2,5-6	rcl
C3	100µF	E2,5-7	rcl
C4	33pF	C0805	rcl
C5	100pF	C0805	rcl
C6	47µ	E2,5-6	rcl
C7	100nF	C0805	rcl
C8	100nF	C0805	rcl
C9	10uF	C0805	rcl
C10	100nF	C0805	rcl
C11	100nF	C0805	rcl
C12	470uF	E3,5-8	rcl
C13	470uF	E3,5-8	rcl
D1	1N4148	DO35-7	diode
F1	6.3A	SHK20L	fuse
F2	6.3A	SHK20L	fuse
K1	BC557B	TO92-EBC	transistor-pnp
K2	BC557B	TO92-EBC	transistor-pnp
K3	BC547B	TO92-CBE	transistor
K4	BC547B	TO92-CBE	transistor
K5	BC557B	TO92-EBC	transistor-pnp
K6	BC557B	TO92-EBC	transistor-pnp
K7	BD139	TO126V	transistor
K8	BC547B	TO92-CBE	transistor
K9	2SA1930	TO220	transistor
K10	2SA1943	TO-264	TO264
K11	2SC5171	TO220	transistor
K12	2SC5200	TO-264	TO264
L1	2.3uH	TJ3-U2	rcl
LSP1	+20V	LSP13	solpad
LSP2	GND	LSP13	solpad
LSP3	-20V	LSP13	solpad

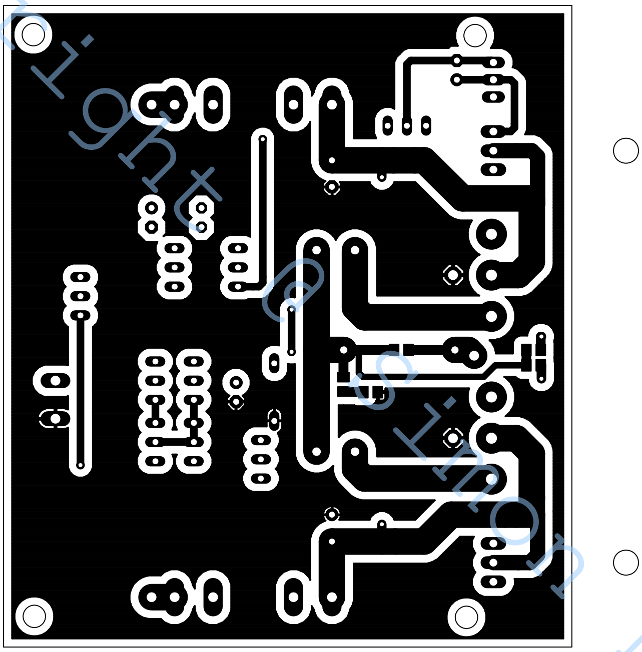
LSP4	OUT	LSP13	solpad
LSP5	GND	LSP13	solpad
P1	Rot	LED5MM	led
R1	39k	R0805	rcl
R2	5k	S64W	pot
R3	39k	R0805	rcl
R4	10k	R0805	rcl
R5	10k	R0805	rcl
R6	100R	R0805	rcl
R7	100R	R0805	rcl
R8	68R	R0805	rcl
R9	68R	R0805	rcl
R10	10k	R0805	rcl
R11	47k	R0805	rcl
R12	330R	R0805	rcl
R13	1.8k	R0805	rcl
R14	150R	R0805	rcl
R15	1.5k	R0805	rcl
R16	100R	R0805	rcl
R17	1k	S64W	pot
R18	68R	R0805	rcl
R19	10k	R0805	rcl
R20	47R	R0805	rcl
R21	0.22R/5W	MNS-5	resistor-power
R22	0.22R/5W	MNS-5	resistor-power
R23	4.7R	R0805	rcl
R24	10R	R0805	rcl
R25	47R	R0805	rcl
X2	IN	AK300/2	con-ptr500

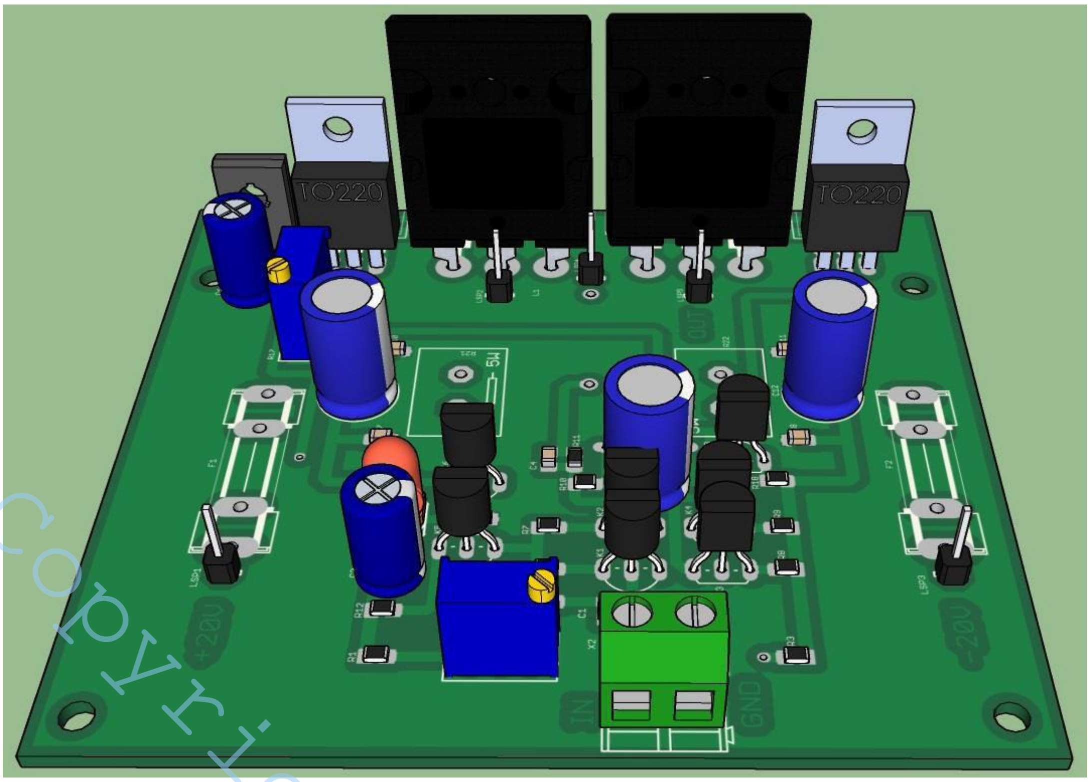
Generated by PART2HTM v0.4 by Sean D. Alcorn
© 1997 Avion International Co. Ltd.
Bangkok • Sydney • Taipei

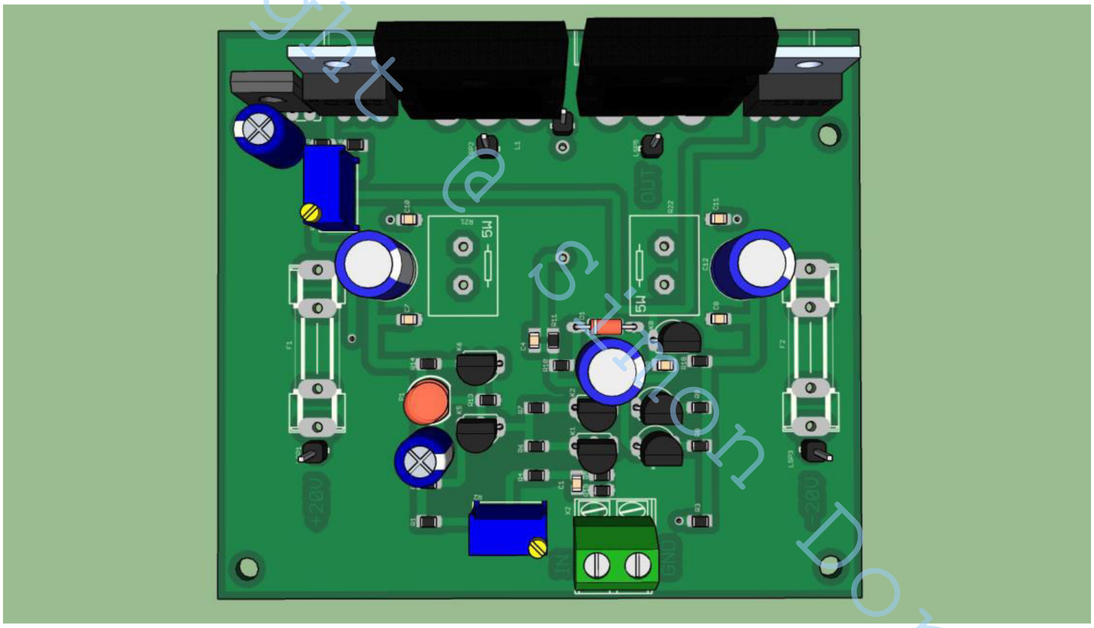

F1
6.3A
F2
6.3A
LSP1
+20V
LSP3
-20V
C13
470uF
C12
470uF
R21
0.22R/5W
R22
0.22R/5W
L1
2.3uH
K11
2SC5171
K12
2SC5200
K10
2SA1943
K9
2SA1930
K7
BD139
IN
OUT
GND




+20V
Dorrer S.
IN
GND
OUT
AB-AMP
-20V




TO220
TO220
5W
+20V
-20V
IN
GND
OUT


5W
5W
+20V
-20V
IN
GND
OUT