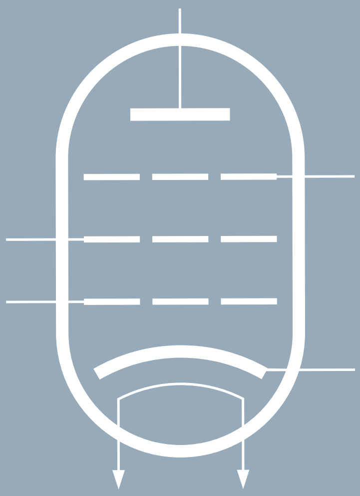
ELEKTOR ACADEMY PRO

High-End Tube Amplifier Design

A Toolbox for Audio Lovers and Engineers



Academy Pro Title by

Zoran M. Dukic

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Contents

To My Daughter Milena

As a focus of the audiophiles, sound is a very complex matter considering its physics and physiology characteristics. Live music sound comes to the audiophile's home audio systems processed by a sophisticated electronics audio chain: live music, recording processes, sound carrier manufacturing. All specifics of human hearing process and differences in physical ambient of live music and home music listening must be taken into account. Most of the problems, including subjective feeling of sound (equiphonic lines or contour), sound levels, frequency response, reverberation time (adopted to law dimension of home listening room), distortion, noise, dynamics (ratio of the highest and the lowest levels of sound in music rendition…) are solved in recording process by very complicated electronic equipment.

Audiophiles typically want to listen the music in the home environment, with sound reproduction which exhibits the highest order of artistic and subjective resemblance to that of the performance in concert hall. At home, music comes from the electronics audio chain (sound currier, reproduction unit, preamplifier, power amplifier and speaker) instead from the music artists. Physical ambient of the home is totally different than the original ambient where the live music is performed. The music artists, audio scientist and audio equipment constructors packed all this things together to fulfill the audiophile's desires, guided by the human ear, with its specific hearing mechanism as the final judge.

Without any ambition to present high level science article, this book has to be accepting as a tool box for audiophiles Hi End equipment designers. The elementary theory is present at the form just necessary to the readers to understand and become familiar with the nature of physics, electric, electromagnetic and electronic process present in certain tool. Each tool is explained in a minimum of words and theory with some equations or figures presented "as is".

So, in the following chapters I will try to guide you through the process of designing quality amplifiers, from the very beginning, fully taking into account both technical and subjective requirements, theory and practice.

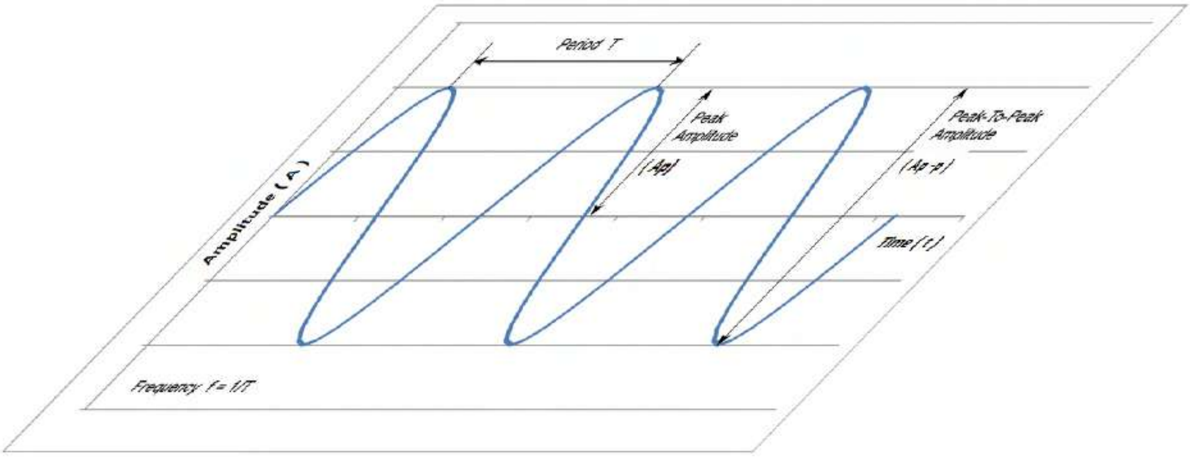
***P**ost **S**criptum*

Some topics are found in unusual places in the book and there is a visible deviation from the methodology of composing a school and science text - the author just wanted to encourage readers on their original path to the realization of their ideas in the world of designing Hi End equipments.

The book is a compilation of the author's notes used in his professional and educational career, but the book was nevertheless primarily written as a result of true love for the audiophile hobby.

Zoran M. Dukic

Chapter 0 • "In Lieu of an Introduction"



I. Units

SI Basic Quantity	Name	Unit
length	meter	m
mass	kilogram	kg
time	second	s
electric current	ampère	A
thermodynamic temperature	kelvin	K
light intensity	candela	cd
matter	mol	mol

SI Electrical Quantity	Unit / Name	Symbol
frequency	hertz	Hz
electrical capacitance	farad	F
force	newton	N
electrical resistance	ohm	Ω
pressure	pascal	Pa
electrical conductivity	siemens	S
energy	joule	J
magnetic flux	weber	Wb
power	watt	W
magnetic flux density	tesla	T
magnetic induction	ampère per meter	A/m
electricity	coulomb	C
inductivity	henry	H
electrical voltage	volt	V
light flux	lumen	lm

SI Electromagnetism Units			
Name	**Symbol**	**Dimension(s)**	**Quantity**
ampère (SI base unit)	**A**	A	Current
coulomb	**C**	$\mathrm{A \cdot s}$	Quantity of electricity
volt	**V**	$\mathrm{J/C = kg \cdot m^{2} \cdot s^{-3} \cdot A^{-1}}$	Potential difference
ohm	**Ω**	$\mathrm{V/A = kg \cdot m^{2} \cdot s^{-3} \cdot A^{-2}}$	Resistance
watt	**W**	$\mathrm{V \cdot A = kg \cdot m^{2} \cdot s^{-3}}$	Electrical power
Farad	**F**	$\mathrm{C/V = kg^{-1} \cdot m^{-2} \cdot A^{2} \cdot s^{4}}$	Capacitance
siemens	**S**	$\mathrm{\Omega^{-1} = kg^{-1} \cdot m^{-2} \cdot s^{3} \cdot A^{2}}$	Conductance
weber	**Wb**	$\mathrm{V \cdot s = kg \cdot m^{2} \cdot s^{-2} \cdot A^{-1}}$	Magnetic flux
tesla	**T**	$\mathrm{Wb/m^{2} = kg \cdot s^{-2} \cdot A^{-1}}$	Magnetic flux density
ampère per meter	**A/m**	$\mathrm{m^{-1} \cdot A}$	Magnetic induction
henry	**H**	$\mathrm{Wb/A = V \cdot s/A = kg \cdot m^{2} \cdot s^{-2} \cdot A^{-2}}$	Inductance
henry per meter	**H/m**	$\mathrm{kg \cdot m \cdot s^{-2} \cdot A^{-2}}$	Permeability

Length	inch **[in]**	foot **[ft]**	yard **[yd]**	**mile**	**mile nautical**	**meter [m]**
inch **[in]**	1	1/12	1/36	0.0156×10^{-3}	0.01371×10^{-3}	25.4×10^{-3}
foot **[ft]**	12	1	1/3	0.1894×10^{-3}	0.1646×10^{-3}	304.8×10^{-3}
yard **[yd]**	36	3	1	0.5683×10^{-3}	0.4937×10^{-3}	0.9144
mile	63354	5278	1760	1	0.8688	1609
mile nautical	72913.38	6076.11	2025.37	1.1510	1	1852
meter **[m]**	39.370	3.2808	1.0936	0.6215×10^{-3}	0.5399×10^{-3}	1

Area	**in^2**	**ft^2**	**yd^2**	**mile^2**	**m^2**
in^2	1	6.94×10^{-3}	0.772×10^{-3}	2.5×10^{-10}	6.4515×10^{-4}
ft^2	144	1	1/9	3.588×10^{-8}	9.2903×10^{-2}
y^2	1296	9	1	3.2294×10^{-7}	0.8361
mile^2	4.01×10^{9}	2785.84×10^{4}	3097.6×10^{3}	1	2.589×10^{6}
m^2	1.5499×10^{3}	10.7639	1.1960	0.3862×10^{-6}	1

Density	**lb/ft^3**	**lb/1n^3**	**kg/m^3**
lb/ft^3	1	5.7875×10^{-4}	16.02
lb/1n^3	1727.84	1	2.768×10^{4}
kg/m3	0.06242	0.3612×10^{-4}	1

Mass	**pound [lb]**	**once [Oz]**	**kilogram [kg]**
pound [lb]	1	16	0.4536
once [Oz]	1/16	1	28.35×10^{-3}
kilogram [kg]	2.2046	35.27	1

Volume	**in^3**	**gallon (USA)**	**gallon (UK)**	**ft^3**	**liter**	**m^3**
in^3	1	0.4323 ×10^{-2}	0.3604 ×10^{-2}	0.5787 ×10^{-3}	16.387 ×10^{-3}	16.387 ×10^{-6}
Gallon (USA)	231	1	0.833	0.1337	3.7854	3.7854 × 10^{-3}
Gallon (UK)	277.42	1.205	1	0.1605	4.5461	4.5461 × 10^{-3}
ft^3	1728	7.4805	6.228	1	28.316	2.8316 × 10^{-2}
liter	61.024	0.2642	0.2199	35.314 ×10^{-3}	1	10^{-3}
m^3	61.024 × 10^3	264.17	219.9	35.3138	10^3	1

Unit multiplies and sub-multiplies

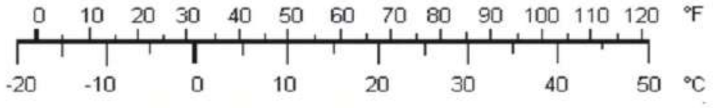
Symbol	**Unit Name**	**Multiply/Divide by**	**Scientific Notation**
T	tera	x 1 000 000 000 000	10^{12}
G	giga	x 1 000 000 000	10^{9}
M	mega	x 1 000 000	10^{6}
k	kilo	x 1 000	10^{3}
m	milli	÷ 1 000	10^{-3}
μ	micro	÷ 1 000 000	10^{-6}
n	nano	÷ 1 000 000 000	10^{-9}
p	pico	÷ 1 000 000 000 000	10^{-12}
f	femto	÷ 1 000 000 000 000 000	10^{-15}
a	alto	÷ 1 000 000 000 000 000 000	10^{-18}

Temperature

$$°\mathrm{C} = \frac{5}{4} \times T_{0_R} = (T_{°\mathrm{F}} - 32) \times \frac{5}{9}$$

$$^0\mathrm{R} = \frac{4}{5} \times T_{°\mathrm{C}} = (T_{°\mathrm{F}} - 32) \times \frac{4}{9}$$

$$°\mathrm{F} = \frac{9}{5} \times T_{°\mathrm{C}} + 32 = \frac{9}{4} \times T_{0_R} + 32$$



II. Some Numerical Values and Mathematics

$$\sqrt{2} = 1.41; \quad \frac{1}{\sqrt{2}} = \mathbf{0.707}; \quad \sqrt{3} = 1.73; \quad \frac{1}{\sqrt{3}} = 0.577$$

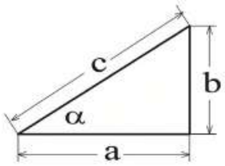
$$\pi = 3.141; \quad 2 \times \pi = \mathbf{6.28}; \quad 1/\pi = 0.318; \quad 1/(2 \times \pi) = 0.159; \quad 2/\pi = \mathbf{0.636};$$

Complex numbers

$$z = a + jb;\ j^2 = -1;\ j = \sqrt{-1};\ (j \times b)^2 = -b^2$$

a = real part; b = imaginary part; j = imaginary unit

Right Angle Triangle



Area: $A = \frac{1}{2} \times a \times b$; Circumference: $O = a + b + c$

Pythagoras theorem: $c^2 = a^2 \times b^2$; $c = \sqrt{a^2 \times b^2}$

Trigonometry formulas:

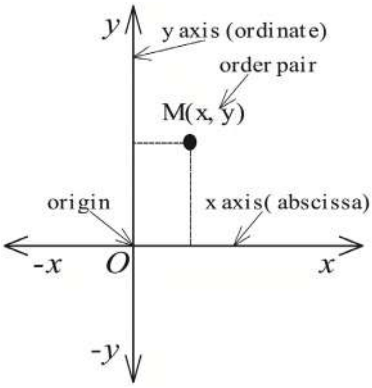
$$\sin\alpha = \frac{b}{c}; \quad \cos\alpha = \frac{a}{c}; \quad \tan\alpha = \frac{b}{a} = \frac{\sin\alpha}{\cos\alpha}\ ; \quad \sin\left(\frac{\pi}{2} - \alpha\right) = \cos\alpha; \quad \cos\left(\frac{\pi}{2} - \alpha\right) = \sin\alpha;$$

Angles (0)	0	30	45	60	90	180	270	360
sin	0	1 / 2	1 / √2	√3 / 2	1	0	-1	0
cos	1	√3 / 2	1 / √2	1 / 2	0	- 1	0	1
tan	0	1 / √3	1	√3	∞	0	∞	0

Complex number – trigonometric form:

$$z = a + jb = r\left(\cos\varphi + j\sin\varphi\right) \qquad r = +\sqrt{a^2 + b^2}; \qquad \tan\varphi = \frac{b}{a};$$

Rectangular coordinate system



A rectangular coordinate system consists of two real number axes that intersect at right angle. The axes intersect at the point O, which is called the *origin.* The horizontal axis is called the *x axis* or *abscissa*, and the vertical axis is called the *y axis* or *ordinate*. These two axes define a flat surface called a plane, and each point on this plane is associated with an ordered pair of real numbers (*x*, *y*) i.e., an ordered pair (*x*, *y*) represents the position of the point relative to the origin.

Mathematical Function

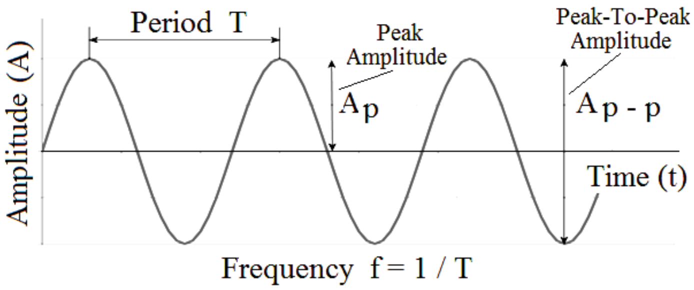
A mathematical function is an expression, rule, or law that defines a relationship between one variable (the independent variable denoted as x) and another variable (the dependent variable denoted as y). This relation is commonly symbolized as *y* = *f*(*x*).

A function can be represented by a table, formula, or graph. Very often in electronics, the representation of some functions in a rectangular coordinate system is used. The independent variable is plotted on the *x* axis and the dependent variable *y* is plotted on the *y* axis. The graph of a function is composed of a series of points with coordinates (*x*, *y*) where *y* = *f*(*x*).

A simplified definition of a mathematical function applicable in electronics is: a special relationship where each input (variable, electrical signal) has a corresponding (depended variable, output signal response or electrical signal).

III. Waveforms

The *Sine* function (Sinusoidal waveform) is a basic function used in analog electronics.



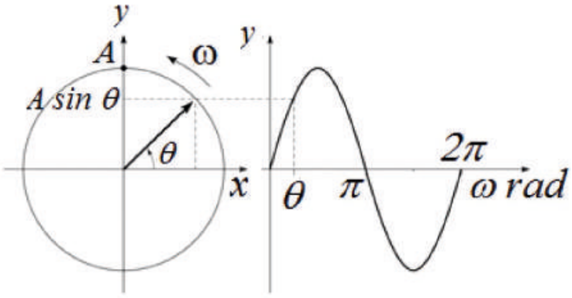
The sine function has a very important application in analog electronics. Many of the variables in analog electronics can be described or expressed by a sine function.

A sine function is a periodic function and in the time domain can be written as:

$$A(t)=A_o \times \sin(\omega \times t)$$

A_0 **–** amplitude (peak or maximum)
ω **–** angular frequency (radians per second)

Another way of representing a sine function (sinusoidal wave form) is in terms of a rotating **phasor.**



A **phasor** has a magnitude ***A*** and rotates at a fixed angular speed ω, so that the angle θ from the ***x***-axis to the phasor increases with time.
The ***y*** - component of the phasor (i.e. its 'projection' ***y = A*** **sin** θ on to the y-axis) varies with θ in the form of a sine function.
The magnitude ***A*** of the **phasor** is equal to the ***amplitude***.
At any time t, the angle θ of the **phasor** is equal to the ***phase*** (generally, ωt + φ).

Radian

A **Radian** (rad) is mathematically defined as a quadrant of a circle where the length of the arc of the circle is equal to the radius (r) of the circle. Since the circumference of a circle is equal to ***2π × radius***, and since a full circle is 360^0 (360^0 = ***2π*** radian), that is **1 radian = 360^0 /2π = 57.30.**

Degrees to radians conversion table:

Degrees	Radians	Degrees	Radians	Degrees	Radians
0°	0	135°	$\frac{3\pi}{4}$	270°	$\frac{3\pi}{2}$
30°	$\frac{\pi}{6}$	150°	$\frac{5\pi}{6}$	300°	$\frac{5\pi}{3}$
45°	$\frac{\pi}{4}$	180°	π	315°	$\frac{7\pi}{4}$
60°	$\frac{\pi}{3}$	210°	$\frac{7\pi}{6}$	330°	$\frac{11\pi}{6}$
90°	$\frac{\pi}{2}$	225°	$\frac{5\pi}{4}$	360°	2π
120°	$\frac{2\pi}{3}$	240°	$\frac{4\pi}{3}$		

If the time of a cycle or period (the time it takes to complete one full cycle) is denoted by T and if one cycles is taken to be equal to 2π radians:

$$\omega=\frac{2\pi}{T} \quad (\text{ radian / s })$$

The reciprocal of the cycle time ***T*** is the **frequency** (cycles per second):

$$\frac{1}{T}=f \quad \text{(unit: Hz)}$$

$$\omega=2\pi\times f=\frac{2\pi}{T}; \quad f=\frac{1}{T}=\frac{\omega}{2\pi}; \quad T=\frac{1}{f}=\frac{2\pi}{\omega};$$

In general:

$$A(t)=A_0\times\cos(\omega t+\varphi)=A_0\times\sin\left(\omega t+\varphi+\frac{\pi}{2}\right)$$

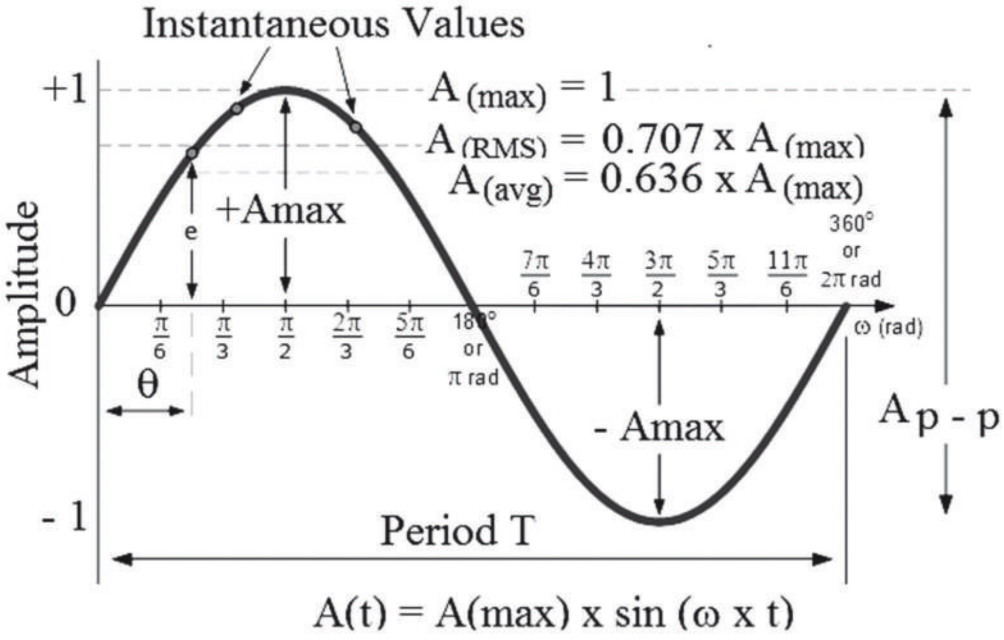
φ *– phase angle*

Other relations:

$$\sin^2\alpha+\cos^2\alpha=1; \qquad \tan\alpha=\frac{\sin\alpha}{\cos\alpha}$$

The sine function describes alternating voltage (U_{ac}) and curent (I_{ac}).

Sinusoidal waveform



Definitions

▲ ***Peak value;* (U_P)**

The peak value is the maximum positive or negative value of an alternating waveform, such as voltage (U_P) or current, during a given time interval.

▲ **Peak–to–peak (p-p) value; (U_{P-P})**

Peak – to – peak value is the difference between the maximum positive and negative values of an alternating wave form in a complete cycle.

$$U_{p-p}=2\times U_p$$

▲ Amplitude

The amplitude is the value of an alternating waveform in the positive or negative direction at the particular moment. The term is often used to denote a peak value.

▲ ***Root – mean – square*** (RMS) value [effective value]; (U_{RMS}), (U_{eff})

$$U_{RMS}=\frac{U_P}{\sqrt{2}}$$

The RMS value, or the so-called effective value, is the square root of the arithmetic mean value of the square of the instantaneous current or voltage values, or other periodic quantities during one complete cycle. The RMS value of a sine wave is the peak value divided by the square root of 2. (Electrically: RMS is equal to the value of the DC current that would produce the same power dissipation in a resistive load).

▲ Average value; (U_{AV})

The average value of an alternating voltage or current is defined as the average of all values of voltage or current during one half-cycle of the waveform.

Technically: The average value of an alternating voltage or current is the value that would be obtained if the original voltage or current were fully rectified. The average value of a sine wave is the peak value multiplied by 2 and divided by π.

$$U_{RMS}=\frac{U_p}{\sqrt{2}} \qquad U_{AV}=\frac{2\times U_p}{\pi}$$

▲ Phase angle

The phase angle of a waveform is the angular difference between two waveforms of the same frequency.

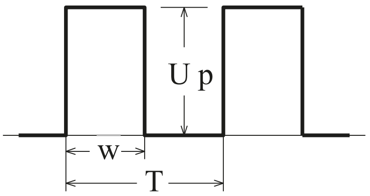
Two waveforms are said to be in phase when they have the same frequency and there is no phase difference between them. (the two waves should reach maximum, minimum and zero values simultaneously at the same time).

- Leading phase angle
  A leading waveform is one that is ahead of a reference wave form of the same frequency.

- Lagging phase angle
  A lagging waveform is one that is behind of a reference wave form of the same frequency.

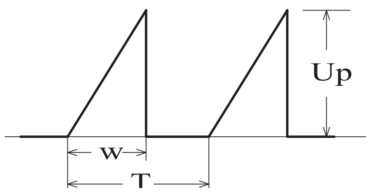
Some types of waveform

RECTANGULAR

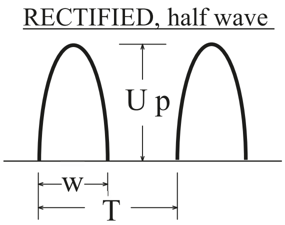


$$U_{RMS}=\frac{U_p\times\sqrt{w}}{T} \qquad U_{AV}=\frac{U_p\times w}{T}$$

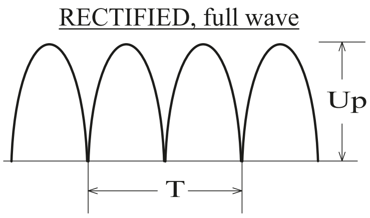
TRIANGULAR



$$U_{RMS}=\frac{U_p\times\sqrt{w}}{3T} \qquad U_{AV}=\frac{U_p\times w}{2\times T}$$

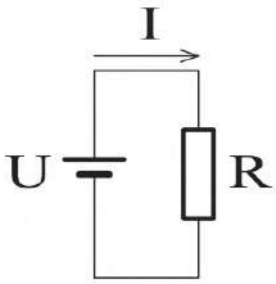


$$U_{RMS} = \frac{U_p}{2} \qquad U_{AV} = \frac{U_p}{\pi}$$



$$U_{RMS} = \frac{U_p}{\sqrt{2}} \qquad U_{AV} = \frac{2\times U_p}{\pi}$$

IV. Ohm’s Law

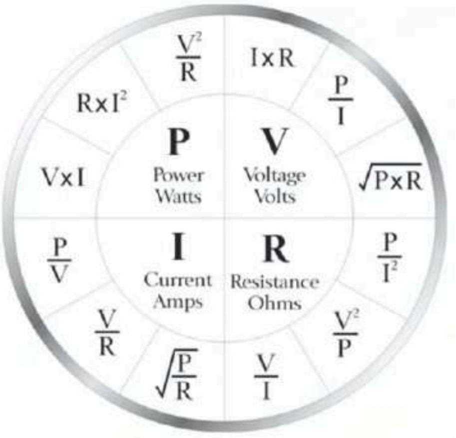


Ohm’s law is an equation that describes the fundamental relationship between voltage, current and resistance, i.e. the law that states that the direct current flowing in a conductor is directly proportional to the potential difference between its ends:

$$I(A) = \frac{U(V)}{R(\Omega)}$$

I=current ; **U**=voltage ; **R**=resistance

I =	$\frac{U}{R}$	$\frac{P}{U}$	$\sqrt{\frac{P}{R}}$
U =	I × R	$\frac{P}{I}$	$\sqrt{P \times R}$
R =	$\frac{U}{I}$	$\frac{U^2}{P}$	$\frac{P}{I^2}$
P =	U × I	$\frac{U^2}{R}$	$R \times I^2$

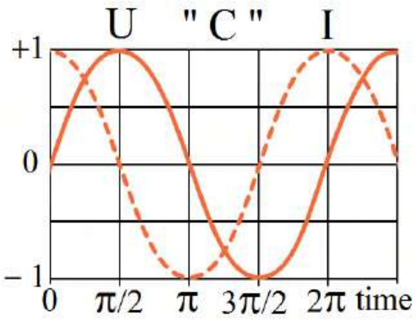


Ohm’s Law is valid for **alternating-current circuits** consisting of **inductance** and **capacitance** as well as resistance:

$$I = \frac{U}{Z} \qquad \frac{1}{Z} = \frac{1}{R} + \frac{1}{jX}$$

Z = impedance

Capacitance, *C*:

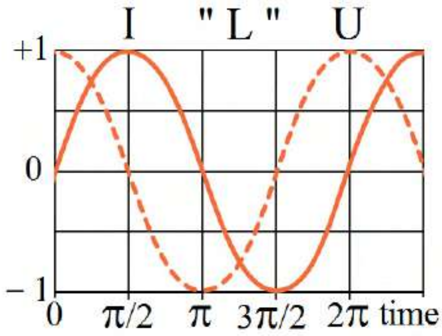


$$Z_c = \frac{1}{j \times \omega \times C} = \frac{1}{j \times 2\pi \times f \times C}$$

$$\frac{U_C}{I_C} = X_C = \frac{1}{\omega \times C} = \frac{1}{2\pi \times f \times C}$$

U lags *I* by π / 2

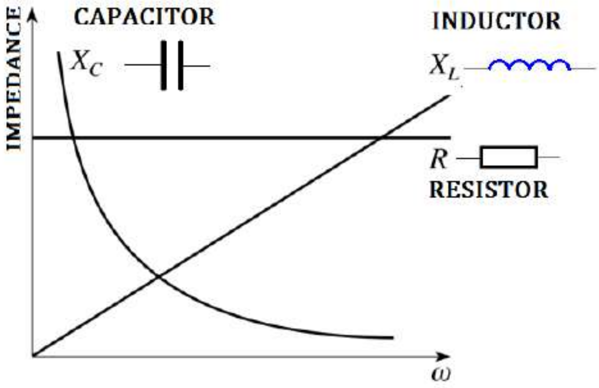
Inductance, L:

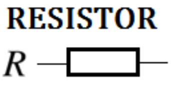


$$Z_L = j \times \omega \times L = j \times 2\pi \times f \times L$$

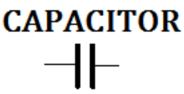
$$\frac{U_L}{I_L} = X_L = \omega \times L = 2\pi \times f \times L$$

U leads *I* by π / 2

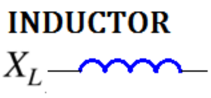




U and *I* in phase



U lags *I* by $\pi/2$



U leads *I* by $\pi/2$

Chapter 1 • Elementary Acoustics

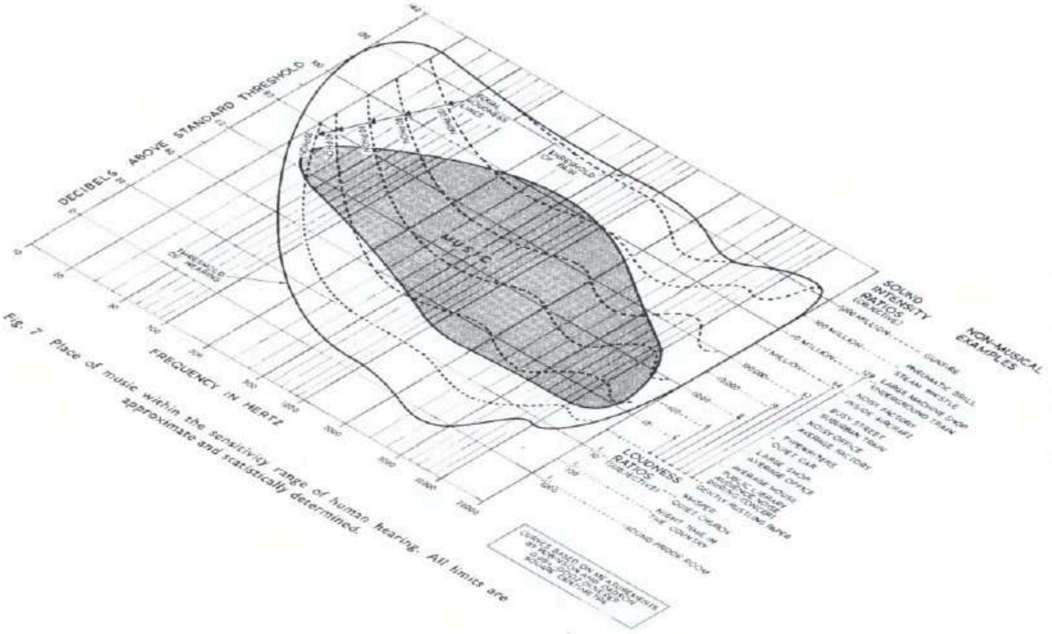
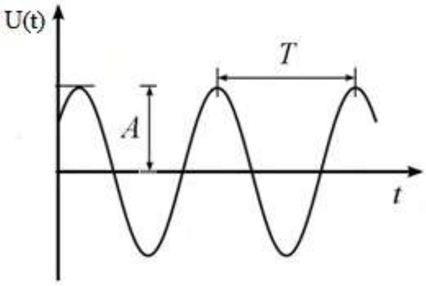


Fig. 7 Place of music within the sensitivity range of human hearing. All limits are approximate and statistically determined.

1.1 Sound

Simplified, sound is defined as **periodic changes in atmospheric pressure caused by the mechanical vibration of the sound source.** From a physiology point of view, only periodic changes in atmospheric pressure in the frequency range between **20 Hz** and **20 kHz,** with a defined intensity, can be recognized by the human ear as sound.

Pure sound has only one periodic component with a defined frequency (fundamental frequency or fundamental harmonic).



The maximum value of the oscillation excursion is the ***Amplitude*** (*A*).
The time taken by a sound wave to complete one cycle (a time span between two consecutive oscillations) is called a ***Period*** (*T*).
The value that is inversely proportional to the Period (*T*) is the ***frequency:***

$$f = \frac{1}{T}$$

Figure 1-01.

Complex sound: a sound that has more than one frequency component (sine wave).

Music tones have components with frequencies equal to the fundamental frequency multiplied by an integer (fundamental and higher-order harmonics).

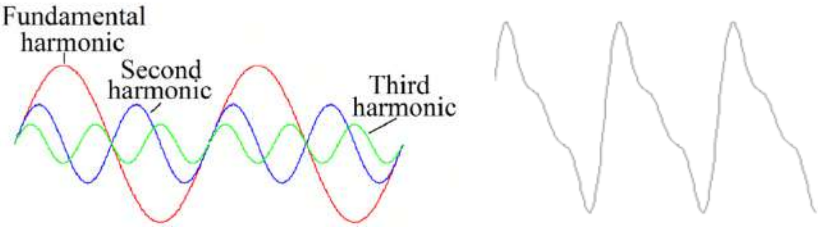


Figure 1-02.

The **distribution** of higher order harmonics and their **intensity** are responsible for the specific **color** of the sound.

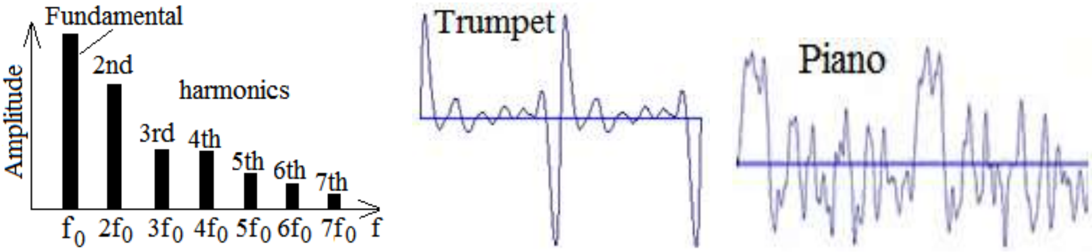


Figure 1-03.

According to physics and physiological research, the sense of sound **loudness** (sound level) basically depends on sound pressure or sound intensity.

The general equation of sound **intensity (J)** valid for all types of sound:

$$\boldsymbol{J} = \frac{p^2}{\rho} \left[\frac{W}{m^2}\right] \tag{1-01}$$

ρ – air density (1.29 kg/m^3)
c – sound velocity in air

($c = 330.7 \times \sqrt{1 + 0.00367 \times t\,[°C]}\ [m/s]$ (sound velocity in air at the temperature *t*)

p – sound pressure (N/m^2)

Pressure units:

$$1\frac{N}{m^2} = 10^{-1}b = 1Pa \tag{1-02}$$

b – bar; ***Pa*** – Pascal

The lowest sound pressure below which the human ear cannot perceive sound (the lowest level of audible sound), is the level of hearing or **threshold of hearing (threshold of audibility)**.

As a result of statistic research, the threshold of hearing (audibility) also known as a **reference sound pressure** at a frequency of 1 kHz is:

$$\boldsymbol{P_0} = 2 \times 10^{-4}\mu b = 2 \times 10^{-5}\frac{N}{m^2} = \mathbf{20}\ \mu\boldsymbol{Pa} \tag{1-03}$$

In practice, it is more suitable to use the units that show the relative relation between two values, known as: **decibels (dB)**:

Logarithm

- Decade: base = 10 — $10^x = N$ — $x = \log_{10}(N) = \log(N)$
- Natural: base = *e* = 2.71828... — $e^x = N$ — $x = \ln N$

log(axb) = log(a) + log(b);
log(a / b) = log(a) – log(b);
log(a^n) = n x log(a);
log(1 / a) = - log(a)

N	**0.001**	0.01	0.1	**1**	10	100	**1000**
log N	**-3**	-2	-1	**0**	1	2	**3**

$$\boldsymbol{L(dB)} = 20 \times log\frac{p}{p_0} = 10 \times log\frac{J}{J_0} \quad \text{(sound pressure)} \tag{1-04}$$

$$\boldsymbol{N(dB)} = 20 \times log\frac{U_2}{U_1} \quad \text{(voltage)} \tag{1-05}$$

$$\boldsymbol{N(dB)} = 20 \times log\frac{I_2}{I_1} \quad \text{(electrical current)} \tag{1-06}$$

$$\boldsymbol{N(dB)} = 10 \times log\frac{P_2}{P_1} \quad \text{(electrical power)} \tag{1-07}$$

The subjective feeling of sound level or loudness is nearly proportional to the logarithm of physical excitations (e.g., of sound pressure) [Weber–Fechner's law].

The level of sound that causes pain in the ear is known as the **level** or **pain threshold** (130 dB).

Conversion to dB of voltage, current and power ratios.

dB	*U, I*	*P*	dB	*U, I*	*P*	dB	*U, I*
0	1.00	1.00	**25**	17.80	316	**50**	316
1	1.12	1.26	**26**	19.90	398	**52**	400
2	1.26	1.59	**27**	22.40	501	**54**	500
3	1.41	2.00	**28**	25.10	632	**56**	632
4	1.60	2.50	**29**	28.20	794	**58**	800
5	1.78	3.16	**30**	31.60	1000	**60**	1000
6	2.00	4.00	**31**	35.50	1260	**62**	1260
7	2.25	5.00	**32**	39.70	1585	**64**	1585
8	2.50	6.30	**33**	44.70	1995	**66**	1995
9	2.82	8.00	**34**	50.10	2515	**68**	2515
10	3.16	10.10	**35**	56.20	3265	**70**	3265
11	3.55	12.60	**36**	63.10	3980	**72**	3980
12	4.00	15.90	**37**	70.80	5010	**74**	5010
13	4.47	20.00	**38**	79.40	6310	**76**	6310
14	5.00	25.10	**39**	89.10	7945	**78**	7945
15	5.60	31.60	**40**	100.00	10000	**80**	10000
16	6.30	39.70	**41**	112.20	12590	**84**	15850
17	7.10	50.10	**42**	125.90	15850	**86**	19950
18	8.00	63.10	**43**	141.25	19950	**88**	25120
19	8.90	79.40	**44**	158.50	25120	**90**	31620
20	10.00	100.00	**45**	177.80	31620	**92**	39810
21	11.25	125.90	**46**	199.50	39810	**94**	50120
22	12.60	158.50	**47**	223.90	50120	**96**	63100
23	14.15	199.50	**48**	251.00	63100	**98**	89125
24	15.70	251.00	**49**	281.80	79430	**100**	100000

Table 1-01.

Sound levels and corresponding sound pressures and sound intensities

Examples	**Sound Pressure Level [dB]**	**Sound Pressure N/m² = Pa**	**Sound Intensity watts/m²**
Pain threshold	**130**	63.2	10
Discomfort threshold	**120**	20	1
Chainsaw, 1m distance	**110**	6.3	0.1
Disco, 1 m from speaker	**100**	2	0.01
Diesel truck (10 m away)	**90**	0.63	0.001
Busy street traffic	**80**	0.2	0.0001
Large office	**70**	0.063	0.00001
Conversational speak, 1m	**60**	0.02	0.000001
Average home	**50**	0.0063	0.0000001
Quiet library	**40**	0.002	0.00000001
House, country	**30**	0.00063	0.000000001
Background in TV studio	**20**	0.0002	0.0000000001
Rustling leaf	**10**	0.000063	0.00000000001
Hearing threshold	**0**	0.00002	0.000000000001

Table 1-02.

The sensitivity range of human hearing is shown in Figure 1-04.

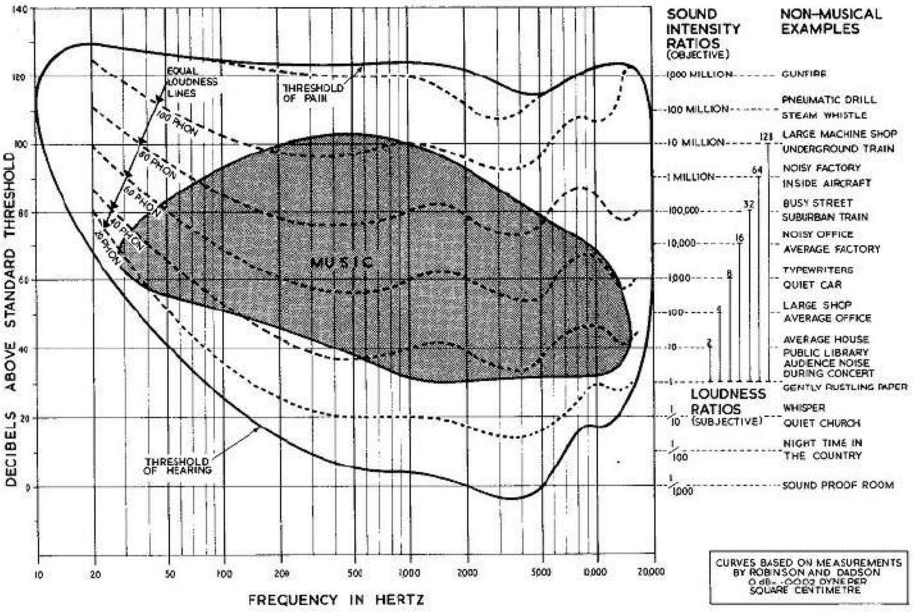


Fig. 7 Place of music within the sensitivity range of human hearing. All limits are approximate and statistically determined.

Fig. 1-04.

By analyzing Figure 1-04, a strong dependence of ear sensitivity on sound frequency can be noticed. Statistical studies of subjective feeling of sound of 1 kHz at a known level and the subjective feeling of sound of various frequencies at equal sound pressure give **equiphonic lines** or lines of equal sound intensity (subjective) in Figure 1-05. But objective measurements that are realized using microphones, audio amplifiers and other measuring instruments, due to their constant sensitivity at all frequencies, always give equal values of sound intensity.

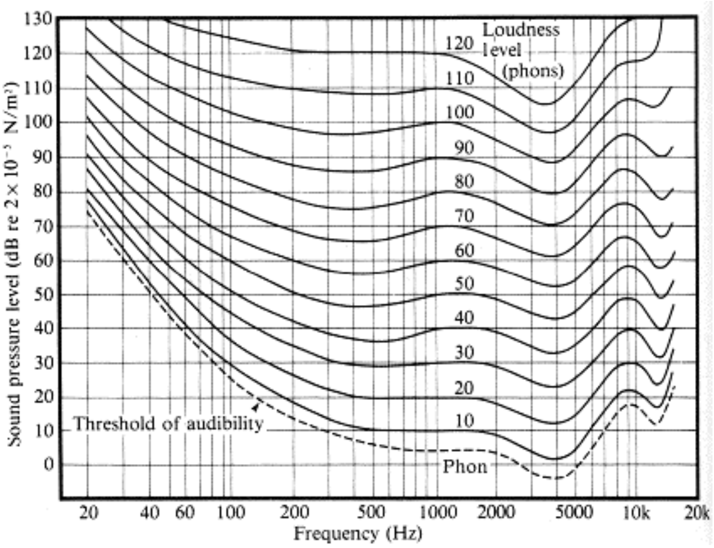


Figure 1-05.

The intensity of sound waves traveling through the air decreases exponentially:

$$J = J_0 \times e^{-(m\times x)} \tag{108}$$

J_o – sound intensity at x = 0
J – sound intensity at distance x
m – constant (depends on temperature and humidity of the air)

If there are some physical barriers in the path of the sound wave, the sound can be **absorbed or reflected**. This is very important for sound propagation in an enclosed space such as a home environment. The incident and reflected sound waves form equal angles to a line perpendicular to the surface of the barrier (as in optics). Part of the sound energy is reflected and part is absorbed in the process of collision of the sound wave with the surface of the barrier. The measure of sound absorption is the **absorption coefficient** α (value between 0 and 1) and depends on the physical characteristics of the barrier material:

$$\alpha = \frac{P_\alpha}{P_{in}} \qquad (1\text{-}09)$$

P_{in} – power of the incident sound wave
P_α – absorbed power

The best sound absorber is an "open window" (α = 1, i.e. total absorption).

The total sound absorption in the room is:

$$A = \alpha_1 \times S_1 + \alpha_2 \times S_2 + \cdots + \alpha_n \times S_n = \sum_{i=1}^{n} \alpha_i \times S_i \qquad (1\text{-}10)$$

S_i – barrier area with absorption coefficient α_i

The total sound absorption in the room is the sum of the sound absorptions by all surfaces: the walls of the room (including the sealing and floor) and other furniture in the room.

In practice, it is more convenient to use the so called **"absorption" (A)** which represents the area of total absorption (α = 1) in relation to sound energy equal to the energy absorbed by the total area covered by the absorption material with the actual value of the absorption coefficient:

$$A = \alpha \times S[m^2] \qquad (1\text{-}11)$$

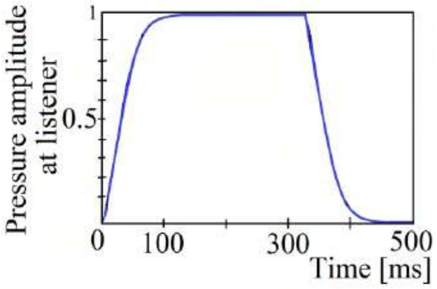


Figure 1-06.

Sound waves that propagate in an enclosed space (room) are constantly in the process of absorption and reflection.

By turning on the sound source in the room, the sound level rises to a stationary value (**J_0**) (incident and absorbed energy in balance). By turning off the sound source in the room, the sound level decreases until it completely disappears (totally absorbed):

$$J = J_0 \times (1 - e^{-(b \times t)}) \text{ (increasing the sound level)} \qquad (1\text{-}12)$$

$$J = J_0 \times e^{-(b \times t)} \text{ (decreasing the sound level)} \qquad (1\text{-}13)$$

b and **J_0** :

$$b = \frac{A \times c}{4 \times V} \qquad (1\text{-}14)$$

$$J_0 = \frac{4 \times P_a}{A} = \frac{p^2}{\rho \times c} \qquad (1\text{-}15)$$

c – velocity of sound
J – sound intensity
V – room volume
P_a – power of the acoustic source
A – absorption

$$P_a = \frac{4 \times \pi \times r^2 \times p^2}{\rho \times c} \tag{1-16}$$

$$J_0 = \frac{25 \times P_a \times T}{V} \tag{1-17}$$

The sound decay time depends on to the acoustic characteristics of the room (dimension, absorption materials). A measure of sound decay time is a **reverberation time** (the time period in which sound level decreases by 60 dB).

$$T = \frac{0.16 \times V}{A} \tag{1-18}$$

$T(s)$ – reverberation time
$V(m^3)$ – room volume
$A(m^2)$ – absorption

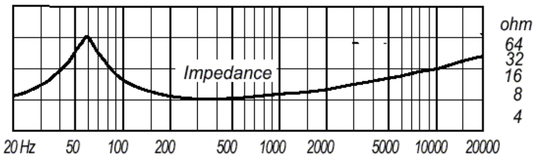
1.2 The Loudspeaker

The loudspeaker converts electrical energy into acoustic energy.

Loudspeaker specifications:

▲ **Nominal (Rated) power or power handling capacity (P)**
declared by the manufacturer as the permissible value of electrical power (in a limited frequency range) applied continuously to the loudspeaker without any possible damage to the loudspeaker.

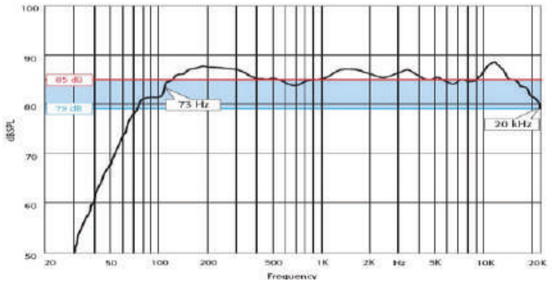
▲ **Impedance (*Z*)**



The impedance is declared by the manufacturer in a limited frequency range. It is ***dynamic impedance***, not DC resistance.

Standardized values: **2Ω, 4Ω, 8 Ω, 16 Ω, 32 Ω, 64 Ω,.800 Ω**.

▲ **Frequency characteristic**



Frequency characteristic (frequency response): produced sound pressure as a function of the frequency ata constant voltage applied to the loudspeaker.

▲ **Sensitivity (e)**

Sensitivity (e): sound pressure at a distance of **1 m** produced by a loudspeaker driven by an electric power of **1 W**.

$$e = \frac{p}{\sqrt{P}} \quad \left[\frac{mb}{\sqrt{W}}\right] \tag{1-19}$$

$$P = \frac{U^2}{Z} \tag{1-20}$$

U – loudspeaker drive voltage (V)
Z – loudspeaker impedance (Ω)

It is common for manufacturers to declare the sensitivity of loudspeaker and loudspeaker boxes in dB:

$$\boldsymbol{P} = 20 \times log\frac{p}{p_0} \tag{1-21}$$

$$\boldsymbol{p} : (P = 1W,\ x = 1m)$$

$$\boldsymbol{p_0} = 2 \times 10^{-4}\mu b$$

Sound pressure decreases with distance i.e. sound pressure is inversely proportional to distance:

$$\boldsymbol{p_x} = \frac{x_1}{x_2} \times p_1 \tag{1-22}$$

$\boldsymbol{p_1}$ – sound pressure at the distance x_1
$\boldsymbol{p_x}$ – sound pressure at the distance x_x

The necessary driving electric power of the loudspeaker to produce the required sound pressure at the distance *x* is:

$$\boldsymbol{P} = \frac{p^2 \times x^2}{e^2} \tag{1-23}$$

p – required sound pressure at the distance x
e – loudspeaker efficiency (the percentage of acoustic energy radiated in all directions from a speaker, compared to a specified amount of driving power (amplifier power)).

▲ Efficiency coefficient (η)

$$\boldsymbol{\eta} = \frac{P_a}{P} \times 100[\%] \tag{1-24}$$

$\boldsymbol{P_a}$ – acoustic power
P – electrical power applied to the loudspeaker

Sound pressure level (J[dB]) produced by a loudspeaker (sensitivity e[dB]) driven by an amplifier P[W], see Table 1-03.

	J [dB]					
	60 soft background music	**70** conversation	**80** Medium volume music	**90** Loud music	**100** Very loud music	**110**
e[dB] loudspeaker sensitivity	**P[W]** amplifier power					
60	1	10	100	1000	10000	100000
70	0.1	1	10	100	1000	10000
75	0.03	0.3	3	30	300	3000
80	0.01	0.1	1	10	100	1000
85	0.003	0.03	0.3	3	30	300
90	0.001	0.01	0.1	1	10	100
95	0.0003	0.003	0.03	0.3	3	30
100	0.0001	0.001	0.01	0.1	1	10

Table 1-03.

1.3 Required output power of an audio amplifier for listening to music at home

According to statistical measurements, the pick level of sound produced by a symphony orchestra in a concert hall is approximately 100 dB. The average listener at home usually listens to music at a sound level of 80 dB. The needs of extreme audiophiles who enjoy listening to music "a little louder" (in the suburb or in the country house) should not be ignored. In order to avoid criticism about the lack of sound level that an audio amplifier should produce for listening to music at home, it is useful to use a sound level of 100 dB as a design requirement in the amplifier design process.

Using equation (1-04), a sound level of 100 dB means that the audio system must produce a sound pressure of:

$$L(dB) = 20 \times \log \frac{p}{p_0}$$

$$100\ dB = 20 \times \log \frac{p}{2 \times 10^{-4}} \quad \rightarrow \quad p = 20 \times 10^{-4}\ \mu b$$

The sound source at home is a loudspeaker driven by an audio amplifier. The electrical power required to drive a loudspeaker of known sensitivity (declared by the manufacturer) and produce a sound pressure of **20×10^{-4}**μb is:

$$e = \frac{p}{\sqrt{P}} \quad \rightarrow \quad P = \frac{p^2}{e^2}$$

The average sensitivity of speakers used for listening to music at home is about **85 dB** i.e. **3.5×10^{-4}** μb:

$$P = \frac{p^2}{e^2} = \frac{(20 \times 10^{-4})^2}{(3.5 \times 10^{-4})^2} = 32.6\ [W]$$

A **2 x 16.3 W** stereo amplifier loaded with **85 dB** sensitivity loudspeakers produces a sound level of **100 dB**. In order to more precisely determine the required electrical power or the output power of the amplifier, the acoustic characteristics of the room must be considered.

The main characteristics of the room that significantly affect the sound level in the room are:

- Room volume
- Area of the walls
- Materials and structure of wall and furniture surfaces
- Room architecture

Using the equations of stationary sound level in the room ($J_0 = 4 \times P_{ac} / A = p^2 / (\rho \times c)$ and absorption [$A = (0.16 \times V) / T$], the equation of acoustic power can be written as follows:

$$\boldsymbol{P_{ac}[W]} = \frac{p^2[\mu b] \times V[m^3]}{T[s]} \times 10^{-6} = 4 \times 10^{\frac{L[dB]}{10}-8} \times \frac{V[m^3]}{T[s]} \times 10^{6}$$

By simplifying the above equation using the empirical architectural equation of area versus room volume:
S = 6.3 x $\mathbf{V^{2/3}}$ and the equation: **A = (0.16** x **V) / T**:

$$\boldsymbol{P_{ac}[W]} = \frac{157 \times \alpha \times 10^{\frac{L[dB]}{10}-8} \times V^{\frac{2}{3}}}{\eta} \times 10^{-6}$$

If, for example, the dimension of an average room for listening to music is: (**5.5** x **6.5** x **2.8) m**, i.e.:
V = 100 m^3
S = 138.7 m^2,

and α = **0.5** (fairly "acoustically dead" room) and the efficiency of the loudspeaker $\eta = 0.01$ (1%), follows:

$$\boldsymbol{P_{ac}}[\mathrm{W}] = \frac{157 \times 0.5 \times 10^{\frac{100[dB]}{10}-8} \times V^{\frac{2}{3}}}{0.01} \times 10^{-6} = \mathbf{16.9\ W}$$

The choice of the output power of the amplifier depends on the required sound level, the efficiency of the loudspeaker and the dimensions and acoustic characteristics of the room for listening to music.

1.4 Conclusion

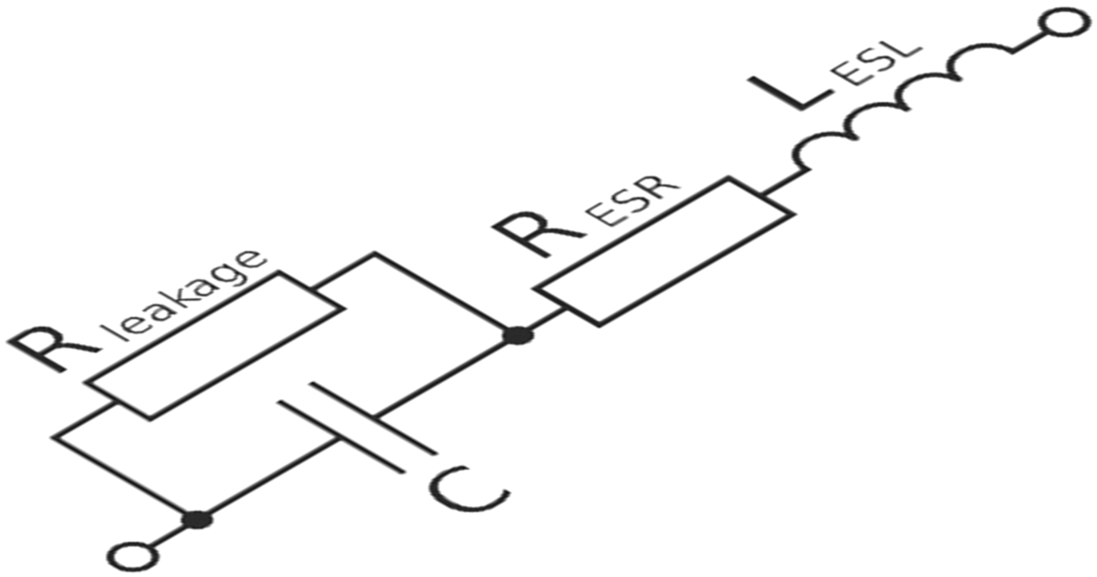
The above analyses (based on the propagation of sound waves in an enclosed space such as a music listening room) show that in an average-sized room a stereo amplifier with an output power of only 2 × 8.5 W, loaded with 85 dB sensitivity loudspeakers, can fulfill the expectations of a music listener regarding the sound level, which is at the level of sound produced by a symphony orchestra in a concert hall. More exact analyzes require detailed characteristics of each material used in the room (absorption coefficients of each material of the room and furniture, exact area and usable volume of the room, etc.,) including the objective auditory characteristics of the music listener.

It is a well-known fact that continuous operation of any electrical equipment at maximum power is not recommended. If the above-mentioned recommendation is applied to the audio power amplifier regarding the output power required for listening to music at home and taking into account the fact that the total harmonic distortion of the amplifier is maximum when the amplifier is operating at maximum power and that it usually increases proportionally to the power at which the amplifier is operating, the empirical recommendation is that the rated (nominal) output power of the home audio amplifier should be at least twice the power calculated by the acoustic analysis above.

High End home audio amplifier output power: 2 × 12 W up to 2 × 30 W.

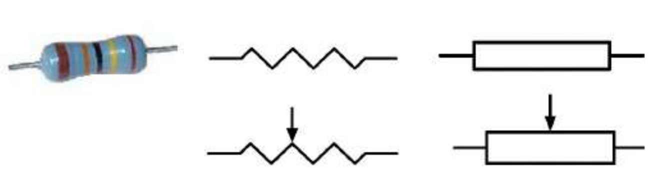
It goes without saying that when using high efficiency loudspeakers, lower output power amplifiers can be used like 2 × 7 W.

Chapter 2 • Passive Electronic Components



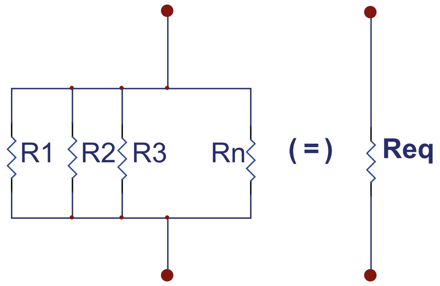
2.1 R – RESISTORS

Unit: **Ω (ohm)** **Fixed resistor:** **Variable resistor:**



A passive electronic component with two terminals that has electrical resistance and is used to control the flow of current in an electronic circuit, i.e. in an electronic circuit a resistor opposes the flow of electrical current through itself. It accomplishes this by absorbing some of the electrical energy applied to it, and then dissipating that energy as heat.

Parallel resistor connection – equivalent resistance:



$$\frac{1}{R_{eq}}=\frac{1}{R_1}+\frac{1}{R_2}+\frac{1}{R_3}+\ldots+\frac{1}{R_n}=\sum_{i=1}^{n}\frac{1}{R_i}$$

Special case – two resistors connected in parallel:

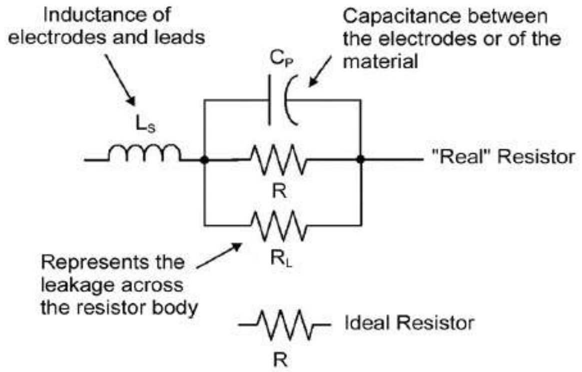
$$\left(R_1 \parallel R_2 \quad R_{eq}=\frac{R_1\times R_2}{R_1+R_2}\right)$$

Series resistor connection - equivalent resistance:

$$R_{eq}=R_1+R_2+R_3+\cdots+R_n=\sum_{i=1}^{n}R_i$$

A model of a real resistor

A real resistor is not just resistance. Some secondary effects caused by the technology and materials used in the manufacturing process appear in the characteristic of the final product - the real resistor:



Inductance: caused by electrodes and leads, geometry of the resistor design, technology and materials used in the manufacturing process.

Capacitance between the electrodes: caused by the electrodes at the ends of resistor and the geometry of the resistor design.

Leakage resistance: caused by the resistance of the body of the resistor**.**

Resistor characteristics:

- Nominal resistance value
  The nominal resistance is standardized and specified by the E-series of nominal standard resistance values.

- Nominal power
  The maximum electrical power permanently applied to a resistor at a specified ambient temperature without degrading its performance.

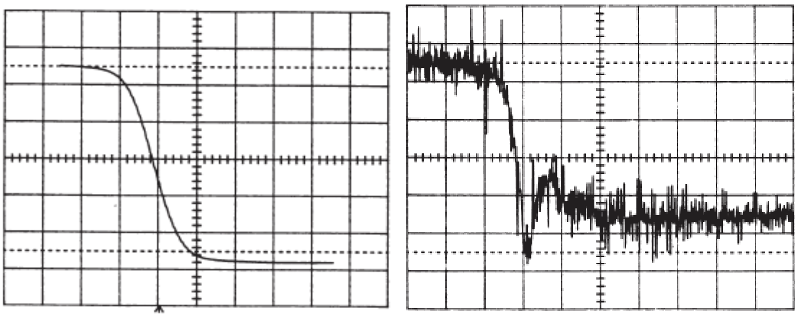
- Operating voltage
  The maximum DC or the maximum RMS AC voltage that can be continuously applied to a resistor at a specified ambient temperature.

- Operating temperature
  The temperature range in which the resistor can be used in continuous operation.

- Temperature coefficient (TK)
  Relative change of resistance per degree of temperature change. It is specified in ppm (part per million, 0.0001%) per degree Celsius of temperature change (ppm / ^{0}C).

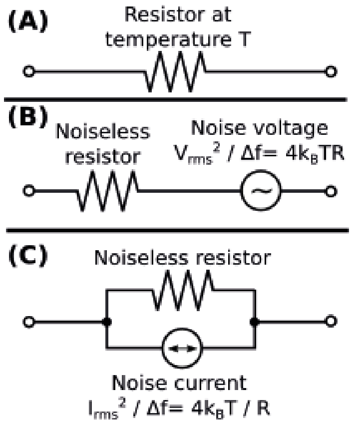
- Tolerance
  The tolerance of the resistor refers to the maximum dispersion of the nominal resistance of the resistor (deviation from the nominal value).

 # Resistor specific characteristic - Resistor noise.



Resistor noise is an unwanted characteristic of a resistor that affects the useful signal and distorts it. It is always present in a resistor and is caused by a fundamental physics process at the electron level.

A real resistor can be modeled by a voltage source (noise voltage source) connected in series with an ideal resistor, or a current source (noise current source) connected in parallel with resistor.



There are two types of resistor noise:
thermal noise and **current** noise.

Thermal noise
Thermal noise is caused by the random motion of electrons inside the resistor material.
The motion of electrons inside the material depends on the temperature and is random.

These motions of electrons produce an electrical signal that is also random (with spectral density uniformly distributed over the frequency range) between the resistor terminals.

The noise voltage (RMS) is:

$$U_{RMS(noise)} = \sqrt{4 \times k \times R \times T \times \Delta f}$$

k – Boltzmann's constant 1.38 x 10^{-23} *J / K*
T – Temperature (K)
R – Resistance (Ω)
Δf – *Frequency range (Hz)*

The thermal noise level is lower if temperature, resistance and frequency bandwidth are lower.
The thermal noise levels of different types of resistors of equal resistance at equal temperatures and at equal frequency bandwidth are equal and are independent of the materials and technology used in resistor manufacturing process.

For example, the thermal noise voltage generated by a 1kΩ resistor at room temperature (300 K) and a frequency bandwidth of 20 kHz is:

$$U_{RMS(noise)} = \sqrt{4 \times k \times R \times T \times \Delta f} = \sqrt{4 \times 1.38 \times 10^{-23} J/K \times 1000\,\Omega \times 300\,K \times 20000\,Hz} = 575\,nV$$

 # Usually, the resistor noise is specified as microvolt noise (µV) per volt (V) of applied voltage, for a 1MHz bandwidth (µV/V)*.*

Current noise

The current noise of the resistor depends on the materials and technology used in the resistor manufacturing process.

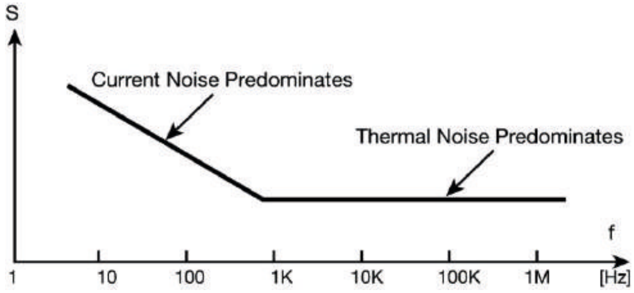
The equation that mathematically expresses the current noise of the resistor is empirical and is usually expressed in ***µV/V*** *or as a* ***Noise Index*** *in decibels:*

$$N(dB) = 20 \times \log\left[\left(u_{(noise)RMS} / U_{DC}\right) \times 10^{6}\right]$$

$u_{(noise)RMS}$ – RMS noise voltage (V)
U_{DC} - DC voltage across the resistor (V)

 # A lower Noise Index means a lower level of current noise of the resistor (the resistor is of better quality).

Diagram of the spectral density of the total noise voltage of the resistor:



Current noise is frequency-dependent and predominates at low frequencies.
The current noise decreases as the frequency increases.

The thermal noise of the resistor is constant over a wide frequency range.

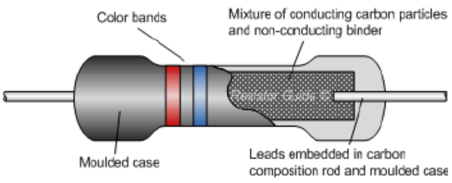
The noise characteristics of the resistors used in the construction of High End audio equipment play a very important role. It is very important to choose low noise resistors for signal path applications as well as power supply chain application.

For example, the input audio signal may be very low, at the level of a few hundred µV (moving coil cartridge of a turntable), and the noise effect of the input resistor interfering with input signal and degenerating it can be very significant, especially since the thus degenerated signal is amplified by next amplifier stages, i.e. the output of the audio amplifier will be distorted and will differ significantly in shape and characteristics from the original input signal. Also, the noise voltage generated in the high-noise resistor applied in the power supply chain of the audio amplifier can interfere with the useful signal and degenerate it.

To improve resistor characteristics and reduce secondary and unwanted effects in resistors such as parasitic capacitance and inductance, leakage current and noise, many different materials and technologies have been developed to manufacture resistors.

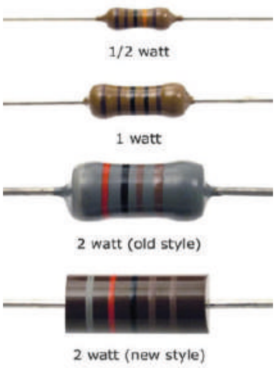
2.1.1 Types of resistors:

- ***Carbon based resistor***
  - Carbon Composition Resistor (CR)



A carbon composition resistor is manufactured from a paste of carbon powder or a mixture of graphite and a non conductive material (such as a ceramic powder) molded and pressed into the desired shape.

Parasitic inductance of a carbon composition resistor is **very low** and they can be used in high frequency circuits.



This type of resistor can be very noisy. The Noise Index is:

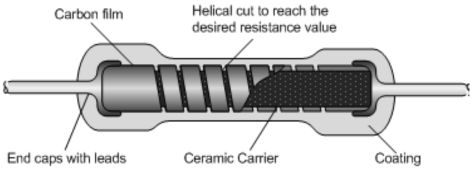
N = (+ 5 dB to - 12 dB).
Temperature stability and Temperature coefficient are not so good and TK can vary in the range from **TK = ± 150 ppm / ^{0}C** to **TK = ± 800 ppm / ^{0}C**.
Power rating: **(1 / 8) W** to **2 W**.

This type of resistor is **not recommended** for use in the construction of High End equipment due to their poor temperature stability, high noise index and high temperature coefficient.

- ***Film resistor***

  *A film resistor is made by depositing a **thick** or **thin** film of a resistive material (carbon, metal (nickel) or metal oxide alloys (tin-oxide or other alloys such as NiCr, tantalum nitride...) on an insulating substrate.*

  - **Carbon film (CFR)**

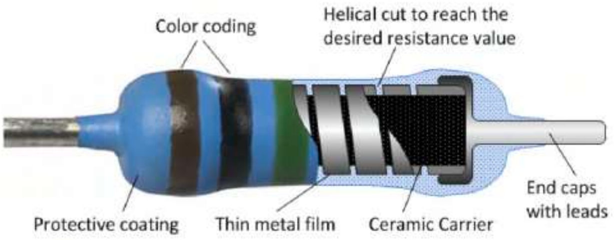


***Carbon** film resistors are an improved versions of resistors made with a carbon base. CFR is manufactured using a thin pure carbon film deposited on a ceramic rod. The deposited carbon film is cut into a helical shape. Unwanted inductance (several μH) and capacitance (around 0.5pF) caused by the helical shape of the carbon film resistive material can affect their operation in RF circuits.*

The Noise Index is better than the Noise Index of carbon composition resistor:

N = (- 5 dB to - 25 dB)
Thermal stability is better than the thermal stability of carbon composition resistor:
TK > ± 250 ppm / ^{0}C
*Maximum operating temperature: up to **200 ^{0}C***
*Power rating: **(1 / 8) W** to **5 W***
 # Application: any

• Metal Film Resistor (MFR) (thin layer)



Thin film resistor is similar in construction to carbon film resistor, but the resistive material is some metal alloy such as Ni-Cr (nickel-chromium), SiCr (silicon-chromium), TaN (tantalum-nitride). MFR is manufactured using a metal film deposited on a ceramic rod. The thickness of the metal film is 10 to 500 Angstroms. The helical shape is made by laser cutting, grinding or chemical etching of the metal film. The unwanted inductance and capacitance caused by the helical shape of the metal film, although significantly smaller than the unwanted inductance and capacitance of carbon film resistors, limit their application in high frequency circuits.

Noise is mainly caused by non-uniform deposition and structural imperfection of the resistive material of the resistor.

The Noise Index is much better than the Noise Index of carbon film resistor: ***N = (- 16 dB to - 32 dB)***
Thermal stability is much better than the thermal stability of carbon film resistor:
TK = (± 5 to ± 50) ppm / ^{0}C

Maximum operating temperature: ***250 ^{0}C to 300 ^{0}C***
Power rating: ***(1/8) W** to **2.5 W***

 # Application: This type of resistor can be used in low- noise circuits.

• Metal Oxide Film Resistor

The construction of this type of resistor is based on a ceramic rod coated with a mixture of tin oxide and antimony oxide.

Metal oxide film resistors can operate at higher temperatures (up to ***450 ^{0}C****)*

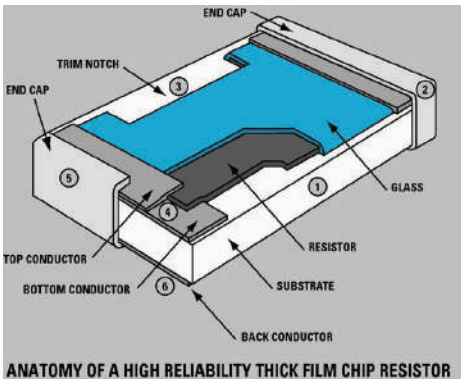
The power rating is higher than the power rating of metal film and carbon film resistors.

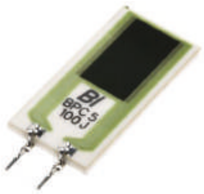
Other characteristics are similar to the characteristics of metal film resistors:

TK = (± 50 to ± 300) ppm / ^{0}C
N = (- 20 to 0) dB

#Application: Highly recommended for use in high voltage and high power circuits, but can also be used in audio amplifier circuits (in the signal path).

• Thick film resistor





The production of thick film resistors is based on the technology of screen printing conductive ceramic paste (such as ruthenium oxide, iridium oxide, rhenium oxide) and glass on a ceramic (alumina Al_2O_3) substrate. The thickness of the film of the resistive material is of the order of 100μm.

Tolerances are usually between 1 % and 5 %, but using laser trimming technology (especially the so-called L cut) tolerances can be extremely low:

≤ *0.1 %. Resistors with such low tolerances can be used in the production of measuring instruments.*

The temperature coefficient is not as good as the temperature coefficient of a thin film resistor:

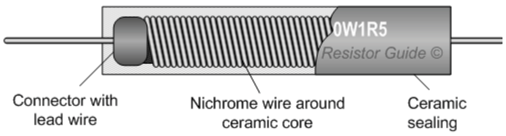
TK =** (**± 50 to ± 200**) **ppm / ^{0}C

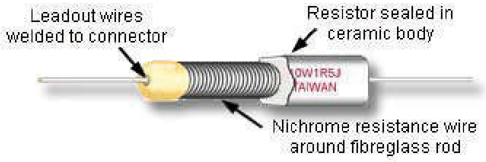
The reason for the not so good Noise Index is the technology used in the production process:
N =** (**- 18 to** - **10**) **dB

This type of resistor is produced in a wide range of power ratings: (1 / 16) W to 250 W.

 # Application: Highly recommended for use in high power circuits, but they can also be used as cathode resistors of high power output tubes (automatic bias).

- Wire wound resistor

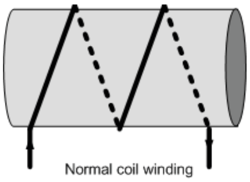


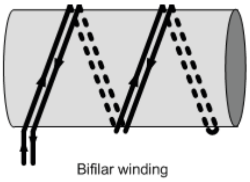


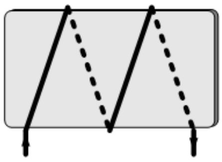
The production of wire wound resistors is based on the technology of winding a conductive wire with a high resistivity around a non-conductive substrate (ceramic rod). Different types of wires are used in the production of resistors, such as alloys of copper, silver, nickel chromium, iron chromium, iron chromium aluminum; marketing names: Constantan, Nickelin, Manganin, Nichrome, Kanthal, Cekas, Megapyr, Tungsten.

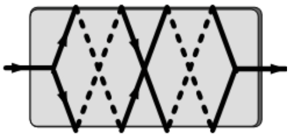
Due to the applied production technology, unwanted parasitic effects such as high inductance and capacitance eliminate this type of resistor for use in high-frequency circuits and in general for use in frequency-dependent circuits.

Several winding technologies have been developed to minimize these effects (bifilar winding, winding on a flat former, Ayrton – Perry winding):


Normal coil winding


Bifilar winding

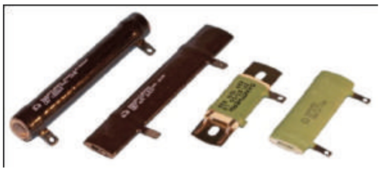

Winding on a flat former


Ayrton-Perry winding

Wire wound resistors have some very good characteristics:

Approaching the lowest current noise compared to other types of resistors.
Noise Index: ***N = - 38 dB (****typical)*
Power rating: ***(1 / 2) W*** *to several* ***kW****.*
Tolerance: ***(± 1 to ± 10) %*** *(typical)*

This type of resistors can be manufactured with a tolerance of the order of ± 0.005 % *and can be used in high accuracy instrument circuits.*

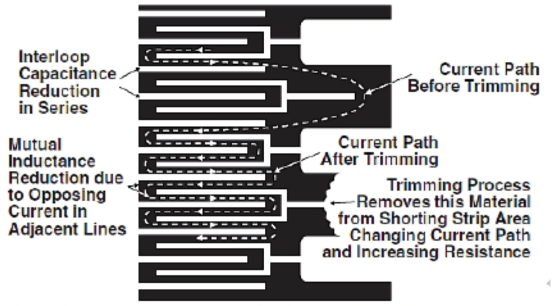


Temperature stability:
TK = (± 10 to ± 200) ppm / ^{0}C
An important property of a wire wound resistor is ***long-term stability:*** *(****15 to 50****) [* ***ppm / year****].*

 # Application: Recommended for use in high power circuits (power supply circuits).
They can be used as cathode resistors (automatic bias).

NOT recommended for use in the signal path** (due to unwanted effects: inductance and capacitance)**.

- Metal Foil Resistor

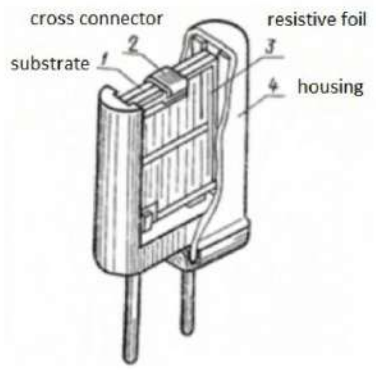


Note: Foil shown in black, etched spaces in white

The manufacturing of metal foil resistors is based on the technology of photo etching of a resistive foil pattern (foil thickness of a few µm) mounted on a ceramic substrate. The foil is usually made of chromium nickel alloys.

The pattern is carefully designed to minimize unwanted parasitic self- inductance and self-capacitance of the resistor:

Inductance: 0.08 µH
Capacitance: 0.5 pF



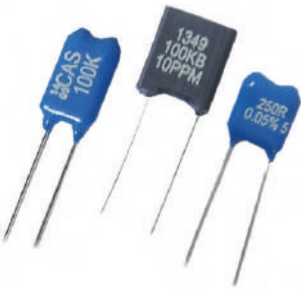
High technology applied in the production process and the laser trimming process result in extremely good characteristic of this type of resistor (already mentioned low inductance and capacitance), such as:

Extremely high precision of the resistance of the resistor.

Tolerance: (± 0.05 to ± 0.001) %
Excellent temperature stability:
TK = (± 0.14 to ± 2) ppm / ^{0}C
Long-term stability is also excellent: ***5 ppm / year***

*The **current noise** of metal foil resistors is **extremely LOW**:*

N = (- 40 to - 42) dB



 # Application: Extremely low parasitic inductance and capacitance, as well as low noise, qualify this type of resistor for use in High End equipment, especially in low level signal circuits such as preamplifier input stage or filter network circuits. They can also be used as active component loads in the signal path.

In general, this type of resistor is the best choice for use in building High End equipment.

Speaking in the language of audiophiles, the use of metal foil resistors in audio equipment results in much cleaner sound, better resolution, lower noise, ...

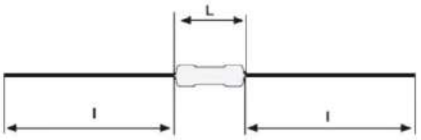
Resistor Type	**Current noise** **µV / V**	**Noise Index** **N [dB]**	**Temperature Coefficient** **TK** [***ppm /*** ^{0}***C]***	**Recommendation** *Application in High End
Carbon composition	0.25 to 1.8	**- 12 to + 5**	± 150 to ± 800	**NO**
Carbon film	0.056 to 0.56	**- 25** to **- 5**	± 250 to ± 800	**YES** *(**NO** in low level signal path)
Thin metal film	0.025 to 0.16	**- 32 to - 16**	± 5 to ± 50	**YES**
Metal oxide	0.1 to 1	**- 20** to **0**	± 50 to ± 300	NO *Power supply circuits, cathode resistors, automatic bias circuits
Thick metal film	0.125 to 320	**- 18 to - 10**	± 50 to ± 200	*Power supply circuits, cathode resistor, automatic bias circuits
Wire wound	0.012	**- 38**	± 10 to ± 200	*Power supply circuits, cathode resistors, automatic bias circuit
Metal foil	0.008 to 0.01	**- 42 to - 40**	± 0.14 to ± 2	**YES**

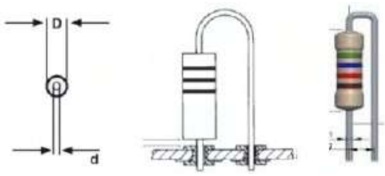
2.1.2 Axial and Radial Resistors

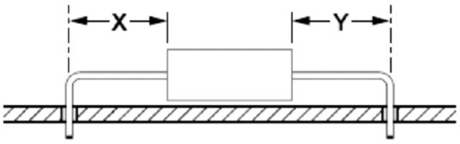
Types of resistors commonly produced and used in practice:

Axial: The leads are located on opposite sides of the resistor body along the body axis.
Radial: The leads are located on the same side of the resistor body.

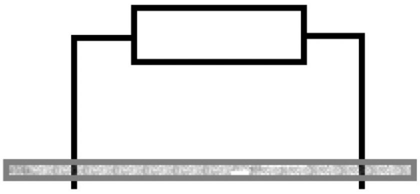
***Axial** resistors are primarily designed for point-to-point wiring, but by bending the leads, axial resistors can be used for PCB mounting* (through-hole mounting)*.*







X is approximately equal to Y

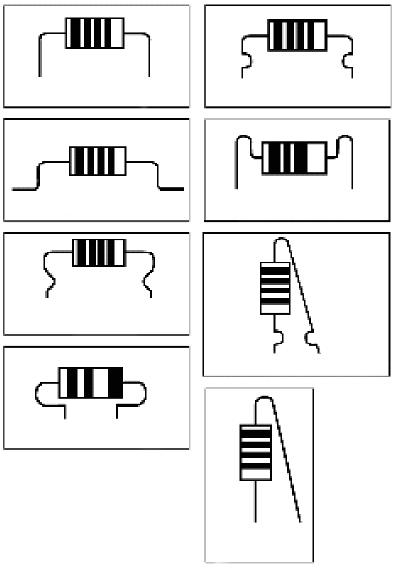


It is common to mount resistors in parallel with the PCB. A small clearance between the resistor body and the PCB is necessary to minimize the vibration transfer between the PCB and the resistor, also the heat radiation of the body of the resistor is more efficient due to the air flow between the body of the resistor and PCB. It is very important to mount the high power resistor at an adequate distance (clearance) from the PCB to prevent extreme heating of the PCB caused by the heat radiated by the high power resistor.

The above precautions must also be taken when mounting resistors vertically on a PCB.

Radial *resistors are primarily designed for PCB mounting, but can be used for point-to-point wiring.*

Bending and forming the leads of axial resistors



2.1.3 Standard Resistor Values

The Electronic Industries Association (EIA) and other authorities specify standard values for resistors.
The EIA “E” series specify resistance values for various tolerances.
The number following the “E” specifies the number of logarithmic steps per decade.

For example, E12 means that every decade (0.1 to 1, 1 to 10, 10 to 100,…) is divided in 12 steps on a logarithmic scale. The size of every step is equal to $10^{(1/12)} \approx 1.21$*. Every resistance value is 1.21 times higher than the previous resistance value in the series, rounded to whole numbers (integer).*

E series (most common tolerances):

E12	10% tolerance,
E24	5% tolerance (also 2% tolerance),
E96	1% tolerance,
E192	(0.5, 0.25, 0.1) % tolerance.

 # Table of the most used E series

E12	E24	E96		E12	E24	E96		E12	E24	E96
100	100	100		220	220	215		470	470	464
		102				221				475
		105				226				487
		107				232				499
	110	110			240	237			510	511
		113				243				523
		115				249				536
		118				255				549
120	120	121		270	270	261		560	560	562
		124				267				576
		127				274				590
		130				280				604
	130	133			300	287			620	619
		137				294				634
		140				301				649
		143				309				665
150	150	147		330	330	316		680	680	681
		150				324				698
		154				332				715
		158				340				732
	160	162			360	348			750	750
		165				357				768
		169				365				787
		174				374				806
180	180	178		390	390	383		820	820	825
		182				392				845
		187				402				866
		191				412				887
	200	196			430	422			910	909
		200				432				931
		205				442				953
		210				453				976

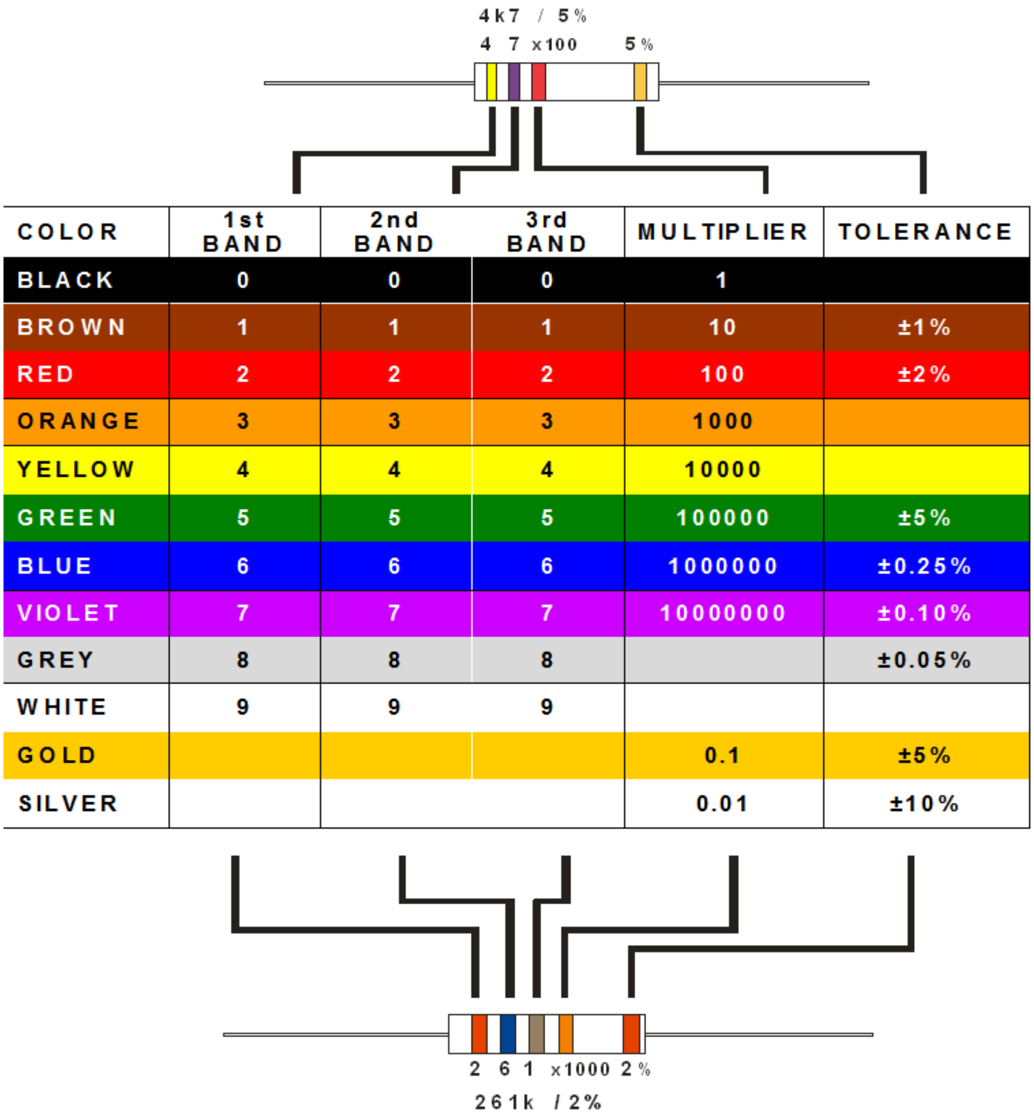
2.1.4 Resistor Color Codes

The values of axial resistors are most often indicated with color codes, i.e. color bands printed on the body of the resistor.

The number of color bands most often used to mark axial resistors is 4 or 5.

- **4 band resistor**
  The first two color bands indicating the significant digits for the resistor value, followed by one color band - the multiplication factor and one color band representing the tolerance

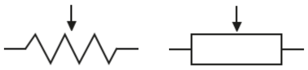
- **5 band resistor**
  The first three color bands indicating the significant digits for the resistor value, followed by one color band - the multiplication factor and one color band representing the tolerance
  Three bands for resistor value, one multiplier and one tolerance band



COLOR	1st BAND	2nd BAND	3rd BAND	MULTIPLIER	TOLERANCE
BLACK	0	0	0	1	
BROWN	1	1	1	10	±1%
RED	2	2	2	100	±2%
ORANGE	3	3	3	1000	
YELLOW	4	4	4	10000	
GREEN	5	5	5	100000	±5%
BLUE	6	6	6	1000000	±0.25%
VIOLET	7	7	7	10000000	±0.10%
GREY	8	8	8		±0.05%
WHITE	9	9	9		
GOLD				0.1	±5%
SILVER				0.01	±10%

The Standard Resistor Color Code Chart

2.1.5 Potentiometers

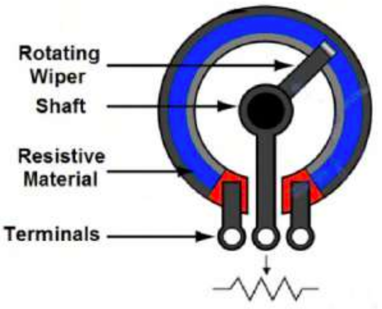


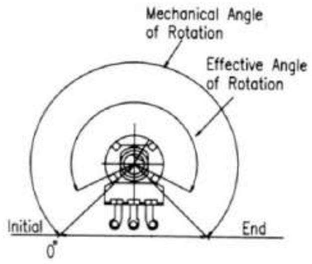
A potentiometer is a manually adjustable variable resistor with 3 terminals. Two terminals are connected to the ends of the resistive element, and the third terminal is connected to a sliding contact, the so-called wiper, which slides over the resistive element.

There are many types of potentiometers:

Rotary - type of potentiometer where the wiper moves along a circular path.
Linear - type of potentiometer where the wiper moves along a linear path.

Construction of the most common type – rotary (single turn):

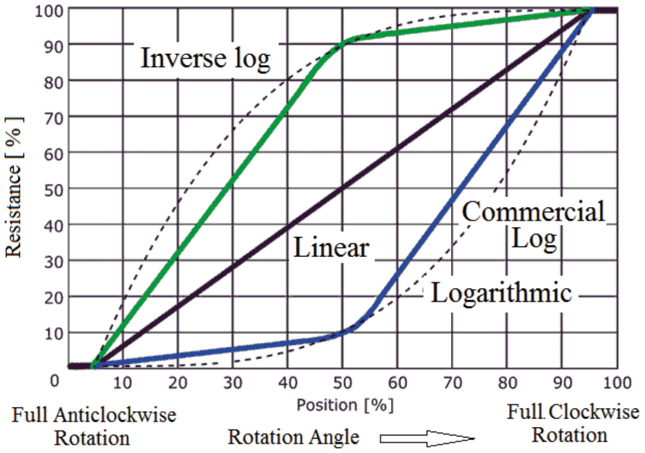




Resistive material:

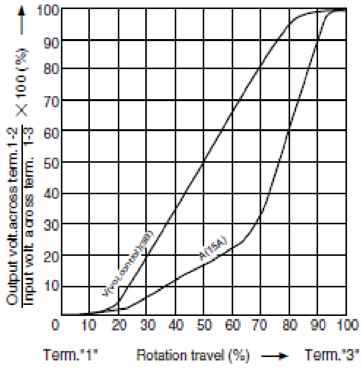
- Carbon composition
- Wire wound
- Conductive plastic
- Cermet (metal film)

One of the most significant characteristics of the potentiometer is the dependence of the resistance between the wiper and one end terminal of the potentiometer on the position of the wiper (R = f (rotation angle) – rotary type and R = f (distance along the linear path) – linear type). A few standard characteristics (percentage of resistance versus angle of rotation) are shown in the following diagram (Linear – Logarithmic (Log) – Inverse log):



For use in audio devices, the so-called **Log** potentiometer (mainly used for audio volume control) is of the greatest interest and is designed so that audiophiles can hear as natural a change in sound level (volume) as possible by rotating the potentiometer shaft. *(The subjective feeling of sound level or loudness is proportional to the logarithm of physical stimuli (e.g. sound pressure) [Weber – Fechner's law]. The sound pressure level is proportional to the output power of the audio amplifier. Perceived loudness varies approximately linearly with the logarithm of the output power of the audio amplifier.*

The resistance characteristic curve of a Log-type potentiometer is basically an exponential curve $R(x) = R_0 (e^{R_1 x} - 1)$ (*x* is the wiper position as a percentage of full scale, R_1 – fitting constant) *i. e. the opposite of the logarithmic behavior of the human ear).*

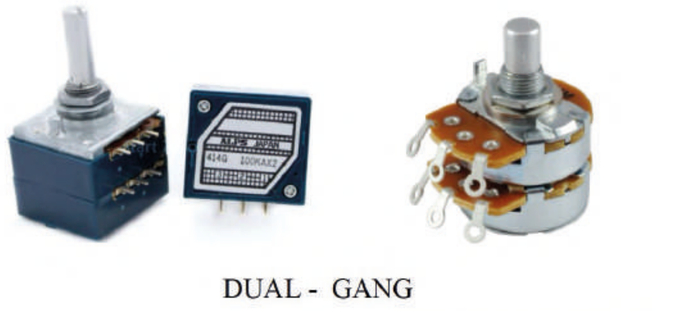


So, the Log potentiometer used in the volume circuit of the power amplifier provides loudness attenuation in dB that varies linearly with wiper position (the loudness changes simultaneously and uniformly as the potentiometer wiper rotates). For example, at a maximum angle of rotation of the Log potentiometer wiper (setting the volume to maximum) the loudness is twice as high as when the wiper of the potentiometer is set to half the maximum angle of rotation (setting to half the maximum volume).

Manufacturers usually produce Log potentiometers with a ratio between wiper position and resistance that is not an exact logarithmic function, but a stepwise approximation of a logarithmic curve (Commercial Log type potentiometer).

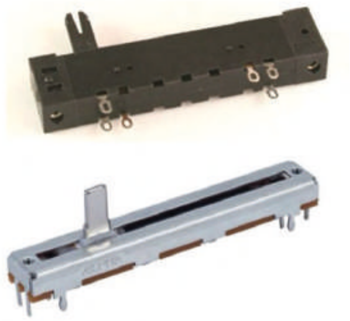
Single - turn rotary Log potentiometers are manufactured as **single** or **dual-gang** potentiometers (*two potentiometer combined on the same shaft – can be used to simultaneously adjust the volume of the left and right channels of the amplifier*).

Mounting style: PCB Mount and Panel Mount (point-to-point wiring).



Specifications
Type: Log
Total resistance: 100 kΩ
Total resistance tolerance: ± 20 %
Max operating voltage: 30 V AC
Rated power: 0.05 W
Total rotational angle: $300^{0} \pm 3^{0}$
Rotational torque: 8 to 35 mN m
Operating life: 15 000 cycles
Operating temperature range: $-10\ ^{0}C$ *to* $70\ ^{0}C$

Note: Consult the manufacturer's Data Sheet on mechanical dimensions



Concentric potentiometer (*two potentiometers, each pot is individually adjusted using concentric shafts)*

Slide pot – linear movement *(single or dual)*

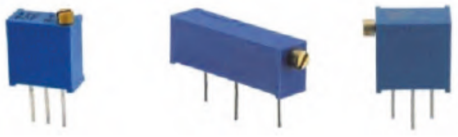
Trimmer potentiometer

Trimmer potentiometers are also known as **preset potentiometers** used for adjustment, calibration and tuning in electronic circuits. They are easily mounted on PCB boards and adjusted with a screwdriver (or small knobs).

Types of trimmer potentiometers: single-turn, multi-turn (usually 10 to 25 turns). Mounting method: through hole, horizontal or vertical mounting. They are produced in a wide range of rated resistance and power.

Single-turn:

Multi-turn:

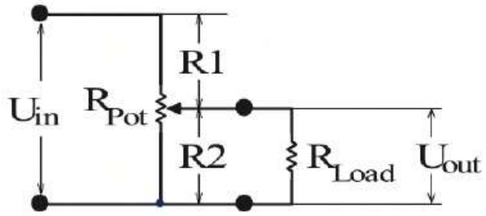


One of the characteristic of a potentiometer is its **electrical noise** (General definition: Any spurious variation in the electrical output not present in the input). The noise in the potentiometer is generated at the contact between the resistive element and the wiper (it depends on the material from which the wiper and the resistive element are made, as well as the force with which the wiper presses the resistive element) – it is measured as a variation of the contact resistance (CRV) expressed a percentage of the initial resistance change and post-test resistance value (usually less than 5%).

In most applications, a potentiometer acts as an adjustable voltage divider (a passive circuit that produces an output voltage that is a fraction of its input voltage).

Example (recommendations):

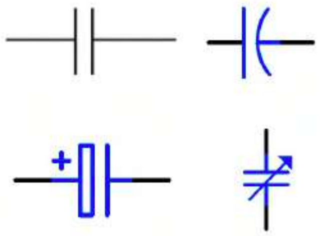
1. It is advisable to choose U_{in} around $U_{in} \approx 2 \times U_{out}$ (or higher).
2. $R_P \leq (<<) R_L$ - (It is advisable to choose a potentiometer with a rated resistance much lower than the load resistance R_L, for example $R_P = R_L / 10$).
3. $P_{P\,(dis)} = (U_{in})^2 / R_P$ - (Always calculate and choose rated power of potentiometer).



$$R_{pot} = R_1 + R_2$$

$$U_{out} = \frac{R_2||R_L}{R_1+(R_2||R_L)} \times U_{in}\ ;\quad R_2||R_L = \frac{R_2 \times R_L}{R_2+R_L}$$

2.2 C – CAPACITORS

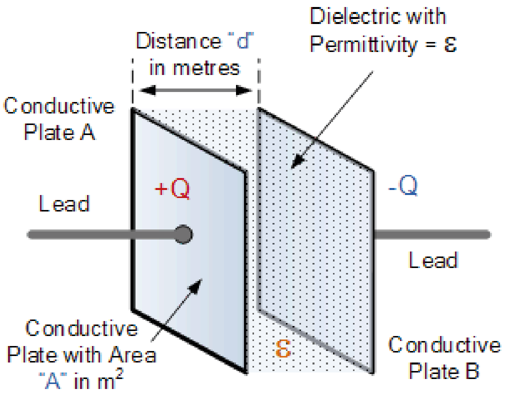


Capacitance C (the amount of charge (Q) that can be stored at a given voltage (U or V) by a capacitor) is the main characteristic of the capacitor:

$$C = \frac{Q}{V}$$

SI unit of capacitance is Farad – [**F**]

- Microfarad (μF) 1μF = 1/1,000,000 = 0.000001 = 10^{-6} F
- nanofarad (nF) 1nF = 1/1,000,000,000 = 0.000000001 = 10^{-9} F
- Picofarad (pF) 1pF = 1/1,000,000,000,000 = 0.000000000001 = 10^{-12} F



A capacitor is a passive electronic component that stores electrical charge or energy in the form of an electrostatic field.

Basically, the capacitor consists of two electrically conductive plates (area of each plate: **S**) separated by an insulating material, the so-called dielectric of dielectric constant **(ε)** or permittivity (the ability of the material to store electrical energy in the presence of an electric field) and thickness **d**.

The dielectric constant of free space (vacuum):

$\varepsilon_o = 8.854 \times 10^{-12}$ F / meter.

Relative permittivity – ε_r, (ratio of the actual permittivity of the dielectric material (ε) to the absolute permittivity of the vacuum (ε_o)).

$$\varepsilon_r = \frac{\varepsilon}{\varepsilon_0}$$, relative permittivity of the air is 1.0006.

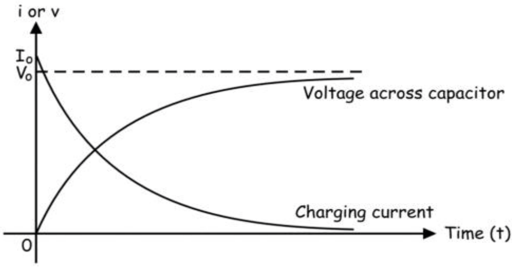
In technical terms: the relative dielectric constant is the ratio of the capacitance of a capacitor with some actual material as a dielectric and the capacitance of a capacitor with a vacuum (or air) as a dielectric.

The capacitance of the capacitor is directly proportional to the area of the conductive plates and the dielectric constant and is inversely proportional to the distance between the conductive plates or the thickness of the dielectric:

$$C = \varepsilon \times \frac{S}{d} = \varepsilon_0 \times \varepsilon_r \times \frac{S}{d}$$, the **basic equation**

When a capacitor is connected to a DC source, the capacitor charges to a voltage equal to the source voltage (fully charged) and then blocks the flow of DC current through it. However, a capacitor in an AC circuit opposes the flow of current through the circuit in a specific way (the capacitor constantly charges and discharges as the polarity of the current changes allowing AC current to flow through the circuit) which is different from the opposition to the current of a resistor (among other things, a capacitor opposes the flow AC currents without heat dissipation. (*ideal capacitor)).*

Capacitor in DC circuit:



Equation of current – voltage of the capacitor:

$$i = C \times \frac{du}{dt}$$

Energy stored in capacitor:

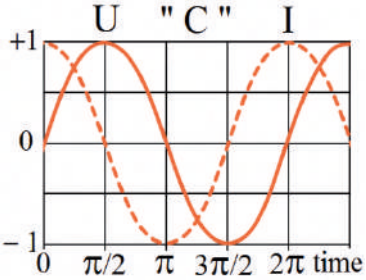
$$E = \frac{1}{2} \times C \times u^2(t)$$

Energy stored in the capacitor at a voltage U:

$$E = \frac{1}{2} \times C \times U^2$$

E [Joules], C [F], U [V] – Voltage across the capacitor

Capacitor in AC circuit:



The ability of capacitor to oppose the flow of AC current (frequency **f**) through the circuit is the so-called **reactance** of capacitor (**X_C**) which is inversely proportional to the frequency of the AC current and the capacitance of the capacitor:

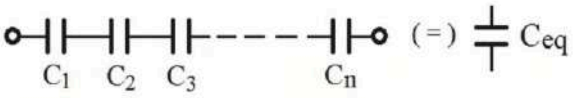
$$X_C = \frac{1}{j \times \omega \times C} = \frac{1}{2 \times \pi \times f \times C}\,; \quad C[F],\ f[Hz],\ X_C\ [\Omega]$$

If the AC voltage $v(t) = V \times \cos(\omega \times t)$ is applied to the capacitor, the current in capacitor is:

$$i(t) = C \times \frac{du}{dt} = -C \times \omega \times U \times \sin(\omega \times t) = C \times \omega \times U \times \cos\left(\omega \times t + \frac{\pi}{2}\right)$$

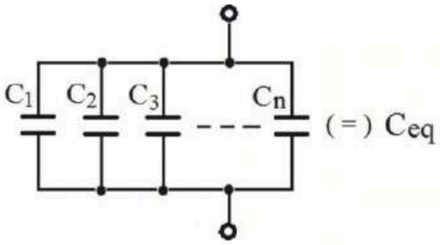
The current **leads** the voltage by $\pi / 2$ in phase. (Voltage **lags** current by $\pi / 2$ in phase).

Series connection:



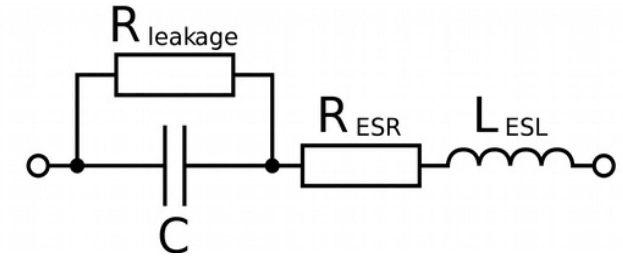
$$\frac{1}{C_{eq}} = \frac{1}{C_1} + \frac{1}{C_2} + \cdots + \frac{1}{C_n} = \sum_{i=1}^{n} \frac{1}{C_i}$$

Parallel connection:



$$C_{eq} = C_1 + C_2 + \cdots + C_n = \sum_{i=1}^{n} C_i$$

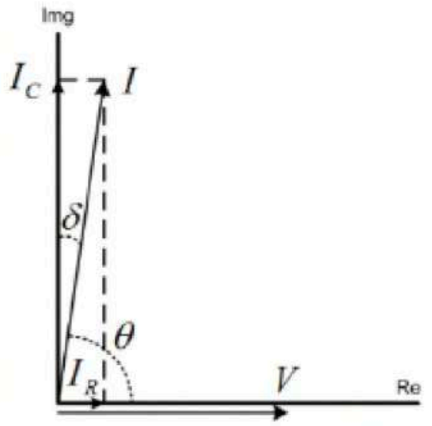
Real capacitor model

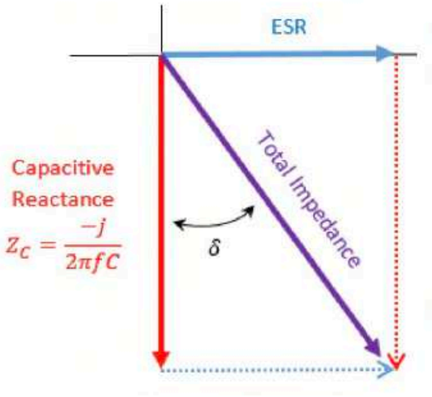


Just as a real resistor is not just resistance, so a real capacitor is not just capacitance. Some secondary effects caused by the technology and materials used in the production process appear in the characteristics of the final product – a real capacitor:

Inductance (L_{ESL}) caused by electrodes and leads, capacitor design geometry, technology and materials used in capacitor manufacturing.

Leakage resistance ($R_{leakage}$) caused by the resistance of the dielectric material**.**





Equivalent Series Resistance (R_{ESR}) caused by wire leads and a characteristic of material used in manufacture of capacitors.

Like any resistor R_{ESR} dissipates energy as heat. A measure of the power loss when an AC signal is applied to a capacitor is the so-called dissipation factor (**DF**) of the capacitor.

Related to the loss factor, in practice the so-called loss tangent (**tan δ**) is more often used (with small differences DF and the loss tangent are the same: DP is used at low frequencies and loss tangent is more suitable for use at high frequencies).

The loss tangent (**tan δ, δ – loss angle**) is defined as the tangent of the difference of the phase angle between capacitor voltage and capacitor current with respect to the theoretical value of 90^0 of an ideal capacitor.

A more general characteristic of the capacitor - The **Quality Factor** (Q) of the capacitor is defined as the ratio of the stored energy of the capacitor to the energy dissipated per cycle. The Q can be expressed the ratio of the capacitive reactance (**X_C**) to the ESR (R_{ESR}) at the frequency f:

$$Q = \frac{X_C}{R_{ESR}}$$

$$tg\delta = DF = \frac{R_{ESR}}{X_C} = \frac{1}{Q}$$

δ - loss angle
DF - dissipation factor
Q - quality factor
ESR - equivalent series resistance
X_c - reactance of the capacitor

The **effective impedance (Z_C)** of the real capacitor is:

$$Z_C = R_{ESR} + j\omega L_{ESL} + \frac{R_{leakage}}{1 + j\omega L_{ESL} R_{leakage}}$$

Capacitor characteristics:

- Nominal value
- Operating voltage
- Operating temperature
- Loss factor (tgδ)
- Temperature coefficient (TK)
- Tolerance

 # Note: Terms used in the application notes for the types of capacitors discussed below:

- ***Coupling capacitor:*** *A capacitor connected in series to the signal path that couples two stages of an amplifier together, allowing AC signals to pass through while blocking DC signals.*

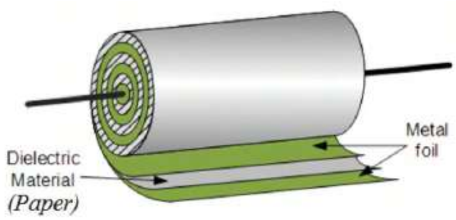
- ***Decoupling capacitor:*** *A capacitor connected in parallel to the power source and the load. It blocks DC signals while allowing AC signals to pass through. It is usually used in a power supply circuit and the power supply line of other circuits. It is used to store electrical energy and remove noise, spikes and ripple from power supply (smoothes out voltage fluctuations).*

- ***Bypass capacitor:*** *A capacitors that short out unwanted AC signals or noise in a circuit. It is usually connected in parallel to a resistor or other electronic or electromechanical component (semiconductors diode, relay contacts, switch contacts,…)*

- ***Energy storage capacitor:*** *A capacitor that can store electrical energy and return it to the circuit. It is usually used in a power supply circuits (filter capacitor). It is used to smooth out voltage fluctuations and remove noise, spikes and ripple from power supply.*

2.2.1 Types of Capacitors

- Paper

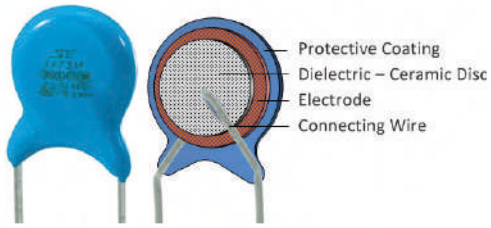


Paper or oil-impregnated (or waxed) paper (dielectric) and aluminum or cooper foil (conductive plates) are used for the production of paper capacitors. They also produced as metalized paper (paper is coated with a thin layer of zinc or aluminum) and paper-in-oil capacitors.

tgδ ≤ 10×10^{-3} to 0.5, TK = ± 800 [ppm / ^{0}C]

 # Application: Recommended for use in audio amplifier circuits (in the signal path – coupling capacitor, bypass capacitor, decoupling capacitor, energy storage capacitor).

- **Ceramic**



tgδ ≤ 2×10^{-3}, TK = (± 30 to ± 2500) ppm / ^{0}C

Depending on the temperature range, temperature drift, and tolerance, ceramic capacitors are classified into the classes:

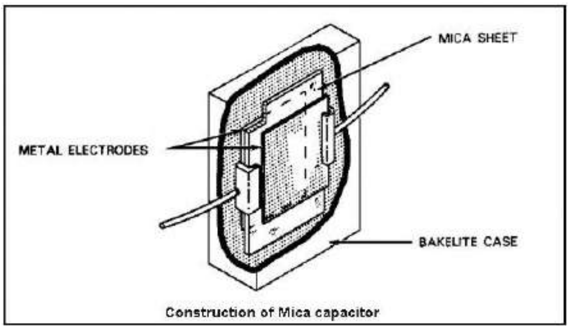
Class (Class 1, Class 2, and Class 3).
Class 1 ceramic capacitors are very stable over a wide temperature range with very linear characteristics.

Capacitance is virtually independent of frequency.

It is most often used in high-frequency circuits (for example, in oscillator circuits or for temperature compensation of frequency discriminating circuits). Can be applied in a very wide frequency band circuit.

 # Application in High End audio amplifier circuit: circuits for compensation and neutralization of parasitic oscillations, filter networks,...

- **Mica**



tgδ ≤ 2.5×10^{-3},

$TK_{typical}$ = + 75 ppm / ^{0}C)
Good temperature stability.
Characteristics stable in time.

 # Application: in a high frequency circuits, filter network circuits, ...

- **Metal – plated plastic (Foil)**

Commonly used in High End equipment (foil can be plated with: **Zn, Al, Cu, Ag, Au**)

Quality of the foil capacitors depended to the dielectric foil characteristic, mainly.

Foil capacitors types:

- **Polystyrene**

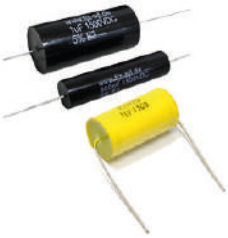
tgδ ≤ 5×10^{-4}, TK = - 125 ± 80 [ppm / ^{0}C]
High insulation
Low leakage
Low dielectric absorption
Low distortion (audio enthusiasts like them because of this)
Good temperature stability
Main disadvantage is low capacitance

Can be used in a signal path – coupling capacitors, high accuracy filter network circuits, frequency compensation circuits, …

The extremely good characteristics of the polystyrene capacitor qualify it for use in almost all High End amplifier and preamplifier circuits .

Application: Recommended for use in any circuit of High End devices.

- **MKP** - polypropylene



tgδ ≤ 10^{-3},
$TK_{typical}$ = + 200 ± 100 [ppm / ^{0}C]
Stable capacitance with time and voltage applied.
Low ESR
Low losses with high current capability, over-voltage capability, …

Application: Coupling, decoupling and bypass capacitor of power amplifier and preamplifier circuits, filter network circuits, power supply circuits - filtering and decoupling, …

- **MKS** - polyester

(tgδ ≤ 5 x 10^{-4}, TK = - 125 ± 80 [ppm / ^{0}C]

Low level ESR, hi and fast signal picks handle dU / dt.

Application in circuits of High End devices: coupling, decoupling and bypass capacitor of power amplifier and preamplifier, filter network, power supply circuits – filtering and decoupling, …

- **MKC, MKM** - polycarbonate

(tgδ ≤ 5 x 10^{-3}, TK ≤ ± 60 [ppm / ^{0}C]

Stable capacitance value with time.
Low dissipation factor.
Stable capacitance value in a wide range of temperatures.

 # Application: power supply circuits - filtering and decoupling, **in a signal path - coupling capacitor, bypass capacitor, …**

Can be replaced with:

PEN - Polyethylene naphtholate
PPS - Polyphenylene sulfide
PI - Polyimide
PTFE - Polytetrafluoroethylene

PTFE or FEP - polytetrafluoroethylene **- Teflon Film**

tgδ = 4 x 10^{-4}, TK = - 200 [ppm / ^{0}C]

Low operating losses.

Low leakage current.
Wide operating temperature range (up to 200^0C), and low temperature drift.
Good stability.

High End circuit application: coupling, decoupling and bypass capacitor in power amplifier and preamplifier, filter network circuits, power supply circuits – filtering and decoupling, ...

The extremely good characteristics qualify this type of capacitor as one of the best choices for use as a coupling capacitor in High End preamplifiers and power amplifiers.

 # Recommended for use in any circuit of High End devices and especially in the signal path - coupling capacitor.

A few more types of foil capacitors:

- **MKT, MKH** - polythereftalate
- **MKU, MKL** - cellulose acetate

Capacitors for special purposes:
A special class of so-called safety capacitors used in the mains supply of electronic devices: Class-X (X1 and X2) and Class-Y (Y1 and Y2) capacitors (also known as EMI/RFI suppression capacitors or AC Line filter safety capacitors).

The X capacitor (usually metalized film capacitors) is connected between the live line (L) and the neutral line (N) of the mains.

The Y capacitors (usually ceramic capacitors, usually used in pairs) are connected across the power lines (L and N) and the ground (G).

The "safety": Safety capacitors are designed to fail in an open-circuit mode or with ability to "self-heal" the dielectric layer (metalized film capacitor,) , and are used in applications where a short-circuit failure would create a risk of fire or electric shock.

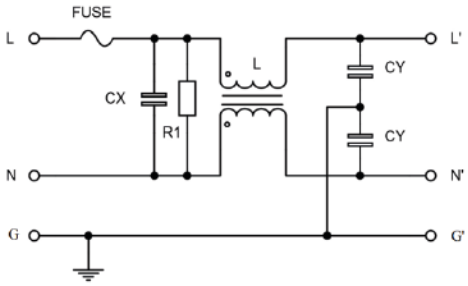
They must be certified according IEC or some other national standards.

Class – X subclass ratings:

Subclass (IEC 60384-14)	Peak Voltage Pulse (while in service)	Peak impulse before endurance test
X1	˃ 2.5 kV ≤ 4.0 kV	4 kV per C ≤ 1 µF 4/√C kV per C > 1 µF
X2	≤ 2.5 kV	2.5 kV per C ≤ 1 µF 2.5/√C kV per C > 1 µF

Class – Y subclass ratings:

Subclass (IEC 60384-14)	Rated Voltage	Peak impulse before endurance test
Y1	≤ 500 VAC	8 kV
Y2	150 VAC ≤ V < 300 VAC	5 kV

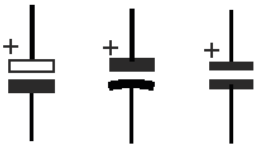


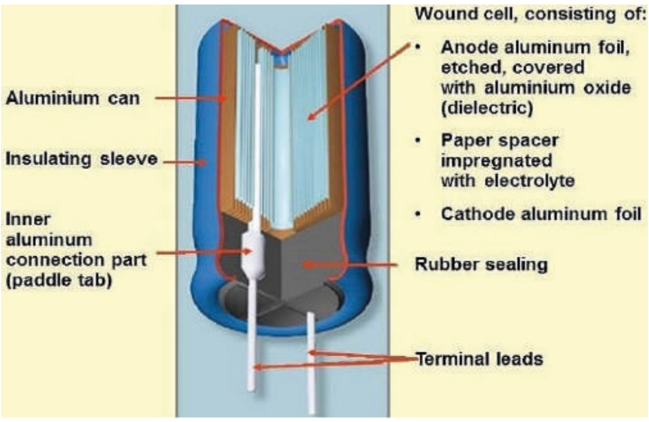
The role of X and Y capacitor when used in an EMI filter: the X capacitor suppresses differential mode interference, and the Y capacitors suppress common mode interference.

For safety reasons, it is extremely important to use certified K and J capacitors manufactured by reputable manufacturers.

- Electrolytic capacitor

Aluminum electrolytic capacitor:





These **polarized** capacitors are made of oxide film on aluminum foils:

cathode aluminum foil, capacitor paper (electrolytic paper), electrolyte, and an aluminum oxide layer, which acts as the dielectric, formed on the anode foil surface.

Due to the specific construction, Al electrolytic capacitors have several advantages compared to other types of capacitors:

- High capacitance (up to few F).
- High ratio of capacitance to volume.

Aluminum electrolytic capacitors are polarized and in working conditions must be connected in a way that ensures their **correct polarization** and operation at voltages within the limits of the nominal voltage.

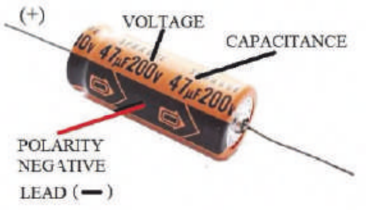
The operation of the capacitor at a lower voltage than the nominal voltage ((recommendation: the value of the working voltage should not exceed the value of 70 ~ 80% of the nominal voltage) increases the life of the capacitor.

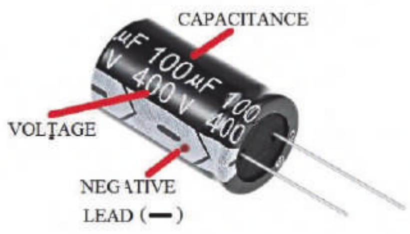
In order to facilitate the identification of capacitors, manufacturers print the basic performance and marking of the capacitors on the bodies of the capacitors (Nominal capacitance [μF], working voltage [V], polarity).

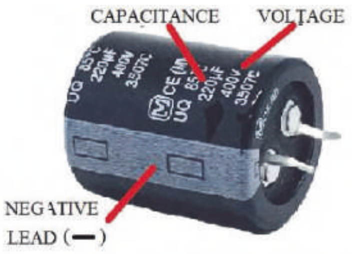
The most common format for marking the polarity of an electrolytic capacitor is to use a stripe on the body of the capacitor. The stripe indicates the negative lead (-).

Some manufacturers practice to indicate the polarity of the capacitor by simply printing a + or - sign on the body of the capacitor.

Axial: the polarity marking strip is usually followed by an arrow pointing to the negative lead.



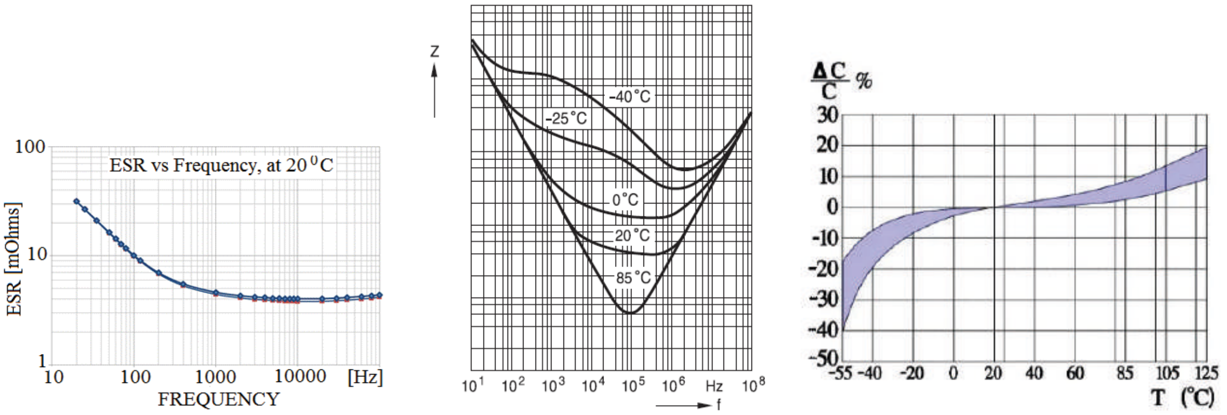




TK = + 2500 [ppm / ^{0}C]

tgδ = (0.1 ÷ 0.28)
ESR and impedance depend on temperature and on frequecy.
High leakage current.
Poor temperature stability.

The working temperature is not high (up to 120^{0}C).



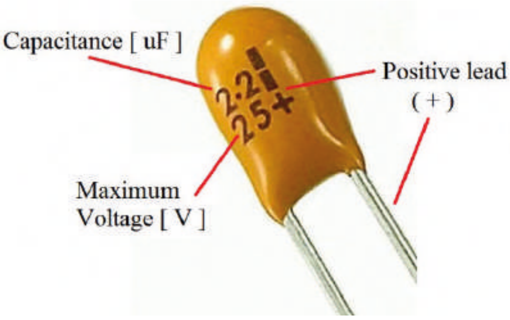
Impedance versus frequency for different temperature values

 # Application: in a power supply filters (smoothing capacitor), decoupling and bypass capacitors, cathode resistors bypass capacitors, *coupling capacitors* ***(****not recommended for use as a coupling capacitor in front of high impedance circuit due the high leakage current****), not recommended for use in high frequency circuits and in wide band circuits.***

However, in any application it is preferable to use aluminum electrolytic capacitors with the lowest possible ESR and tg(δ) - DF (lower heat dissipation, lower harmonic distortion, better frequency response,...)

- **Tantalum**

Type of polarized capacitor.
Tantalum capacitors are made of **metal tantalum** (foil tantalum or sintered tantalum powder) as the anode material.



tgδ = 0.1 to 0.2, TK = + 800 [ppm / ^{0}C]
Large capacitance per unit volume.
Good temperature stability.
Stable over time.
Intolerant to being reverse biased.
Intolerant to high ripple currents or voltages above their working voltage.

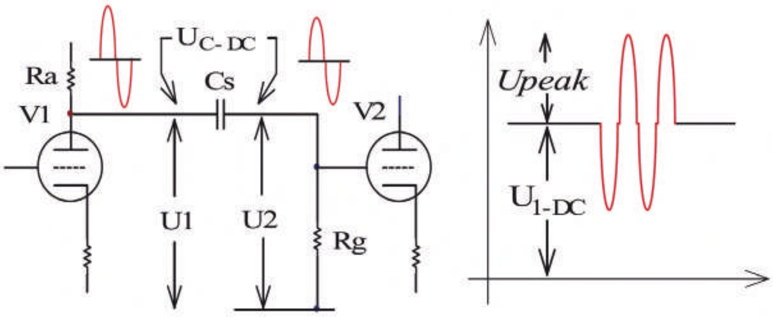
Summary
Table of capacitor types depending on the properties of the dielectric materials used for their production:

Material	Dielectric Constant	Dissipation Factor - tgδ	Temp Coefficient / °C	Notes
Vacuum / Air	1.0 / 1.0006	0	0	Ideal capacitor
Teflon	2.0	0.0004	–200 ppm	high end audio, any circuit, **coupling cap**.
Polystyrene	2.5	0.0005	–125 ppm	high end audio, any circuit, **coupling cap**, **filter network** circuit
Polypropylene	2.3	0.001	-200 ± 120 ppm	high end audio, any circuit, **coupling cap**.
Polyester	3.2	0.0005	-125 ± 85 ppm	general purpose, **decoupling**
Polycarbonate	3.0	0.005	± 60 ppm	decoupling
Paper/wax Paper/oil	2.5 - 4.0	0.02	± 800 ppm	high end audio, any circuit, **coupling cap**.
Mica	5 - 9	0.001- 0.0025	75 ppm	high frequency circuit, filter network circuit
Ceramic	75 typ.	0.002	± 30 ppm	Compensation cap., neutralization of parasitic oscillations, filter networks,…
Aluminum electrolytic		0.1 ÷ 0.28	2500 ppm	**power supply smoothing cap, decoupling, bypass**
Teflon		0.1	800 ppm	

Example
Designers of high-end amplifiers pay the most attention to the coupling capacitor as one of the main electronic components on which the electrical and auditory performance of the audio amplifier depends.

It is very important to carefully carry out the procedure of determining and choosing the coupling capacitor:

1. Nominal voltage



Using the schematic diagram of the example:

a) The value of the working DC voltage on the coupling capacitor is:
$U_{C-DC} = U_1 - U_2$,

b) The alternating signal voltage of the peak value U_{peak} must be considered.

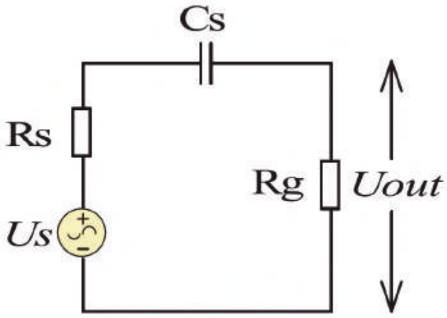
c) The operating voltage on the coupling capacitor is: $U_{C-operating} = U_{C-DC} + U_{peak}$.

d) The nominal voltage of the coupling capacitor must be higher than the operating voltage:
$U_{C-nominal} > U_{C-operating}$

Considering the tolerances of the capacitor and for the safe operation of the capacitor, it is recommended that the nominal voltage of the selected coupling capacitor is 30 to 50% higher than the calculated value of the operating voltage of the capacitor:

$$U_{C\text{ – nominal}} \geq (1.3 \text{ to } 1.5) \times U_{C\text{ – operating}}$$

2. *Nominal capacitance*
   Equivalent circuit at low frequencies:



R_S - *Output resistance of the amplifier stage (tube V 1)*

(For example:

$$R_S = R_a \, II \, R_i = \frac{R_a \times R_i}{R_a + R_i},$$

R_a *– resistance of the anode resistor,* R_i *– internal resistance of the tube).*

R_g *– Input resistance of the other amplifier stage (tube V 2).*
(At low frequencies, the input impedance is equal to the resistance of*the grid resistor).*

By analyzing the equivalent circuit of the RC coupled amplifier, it is observed that the frequency response of the amplifier depends on the output resistance of the input amplifier that acts as a signal source, the coupling capacitor and the input resistance of the next amplifier stage. The frequency response at low frequencies is highly dependent on the capacitance of the coupling capacitor, so it is extremely important to correctly determine its capacitance.

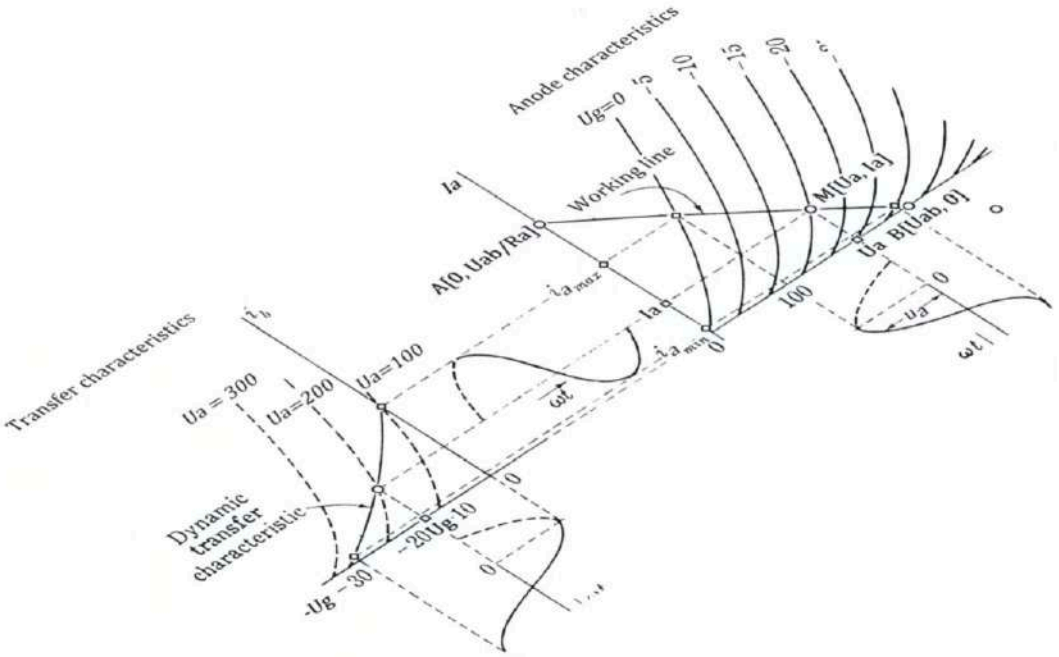
Capacitance of the coupled capacitor for the required -3dB low frequency cut-off is:

$$C_S = \frac{1}{2 \times \pi \times f_{(-3\text{dB})} \times \left(R_S + R_g\right)}$$

3. *Based on the calculated values of nominal voltage and nominal capacitance,* ***standard*** *values are determined that must be higher than the calculated values.*

4. *The process of determining and choosing the coupling capacitor ends with the selection of the type of capacitor depending on the dielectric material:* ***MKP*** *- polypropylene,* ***Teflon, Paper-in-oil****, …*
   The sonic performance of an amplifier depends to a large extent on the quality of the coupling capacitors, so it is necessary to select high-quality capacitors intended for use in audio amplifiers.

Chapter 3 • Fundamentals of Vacuum Tubes



3.1 VACUUM TUBES (BASICs)

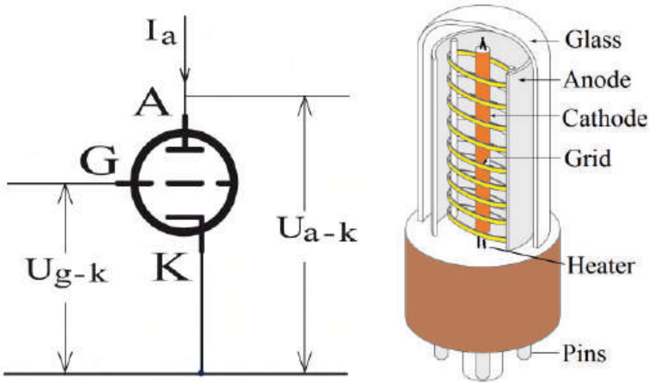


Fig. 3-01

A vacuum tube (valve, electron tube) is an electronic component whose operation is based on the flow and control of the flow of electrons (electric current) caused by the electric field created by electrodes at different electric potentials in a high vacuum.

Triode is a vacuum tube consisting of three electrodes (cathode, grid and anode (plate)) placed inside a sealed highly vacuumed (10^{-3} – 10^{-8} mmHg) glass tube.

CATHODE is one of the electrodes of the vacuum tube that acts as a source of electrons during the physical process of so-called thermionic emission (when the cathode, made of a special material suitable for the thermionic emission, is heated, electrons leave the surface of the cathode).

- Directly heated cathode
  The heater (filament), uncoated or coated with a material suitable for the thermionic emission, is connected to the source of electric power supply and at the same time operates as a cathode.

- Indirectly heated cathode
  The heater is physically (electrically) separated from the material that emits electrons, i.e. the heater coated with ceramic (Al_2O_3) which acts as an insulator, is inserted inside a metal (usually Ni) tube coated with the material suitable for the thermionic emission (usually BaO + SrO).

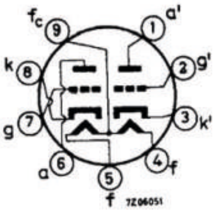
ANODE (PLATE) is the electrode of a vacuum tube that collects electrons emitted by the cathode. The anode is at a positive potential relative to cathode, so, there is an electric field in the space between the cathode and the anode. Due to the action of the electric field, the electrons emitted by the cathode flow towards the anode and hit it, i.e. electric current (anode current) flows through the tube.

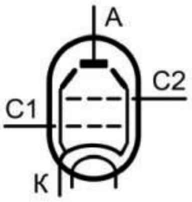
The anode is made of sheet metal most often blackened by an electrochemical and thermal process for more efficient radiation of thermal energy produced by the impact of electrons in the anode (related to the characteristic of the vacuum tube, so-called **anode dissipation**, $\mathbf{P_d = U_{a-k} \times I_a}$)) and to reduce the unwanted effect of the so-called secondary electron emission.

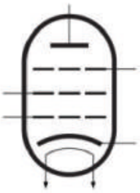
GRID is the electrode of a vacuum tube made of a thin wire located in a space between the cathode and the anode. The value of the current flowing through the vacuum tube (anode current) depends on the distribution of electric field inside the electrode system of the vacuum tube. The distribution of the electric field can be changed by changing the voltage between the grid and the cathode at constant anode voltage, i.e. the anode current is controlled by changing the voltage between the grid and the cathode at a constant anode voltage.

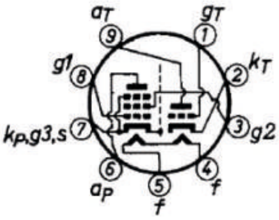
The anode current depends on the anode voltage and the grid voltage, i.e. $\mathbf{I_a = f\,(U_a, U_g)}$. The effect of the grid voltage on the anode current is more significant than the effect of the anode voltage because in the geometry of the electrode system of the vacuum tube, the grid is closer to the cathode than to the anode.

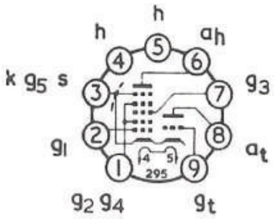
In one tube there can be more than one grid: **triode, tetrode, pentode, hexode, heptode**, Also, in one glass balloon tube there can be more than one identical or different tube system: **double triode, triode pentode,** …

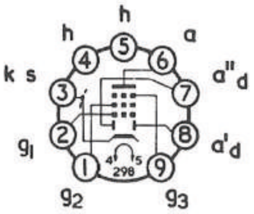


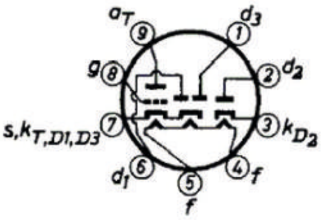


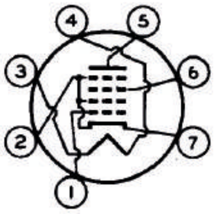






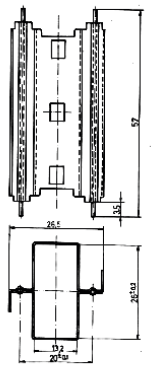




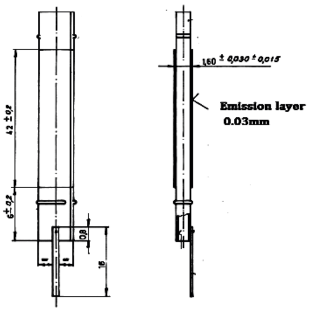


The physical dimensions and shapes of the cathode, grid and anode and their relative distances from each other (the geometry of the electrode system depends on the construction of the tube, i.e. the **type** of tube) determine the distribution of the electric field inside the tube, i.e. static and dynamic characteristics of the vacuum tube.

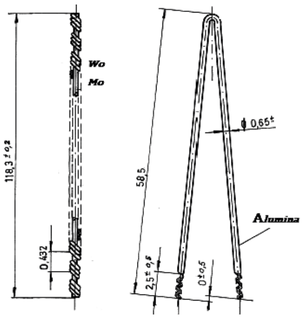
The production of vacuum tubes is very complicated and complex and combines almost all known technological processes – mechanical, chemical, electrochemical, electrical, thermal, vacuum, … A large number of different raw materials such as wires, metal sheets, gases, chemicals, and even water must be of the highest level of purity. Precisely manufactured metal parts (tolerances of the order of microns - anode, heat radiation wings, supporters. ,,,) are treated by chemical cleaning and thermal treatment.



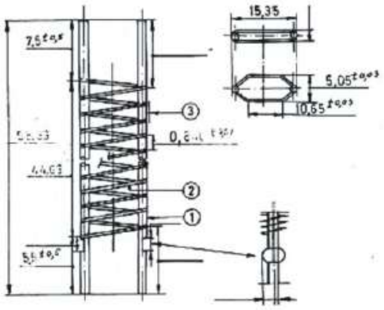
The anodes are made of a specific multi-layer metal sheet:
Ni Fe Al or Al Fe Cu Fe Al.



The cathodes are made of Ni tubes or Ni sheets and are coated with emission suspension (BaO SrO).

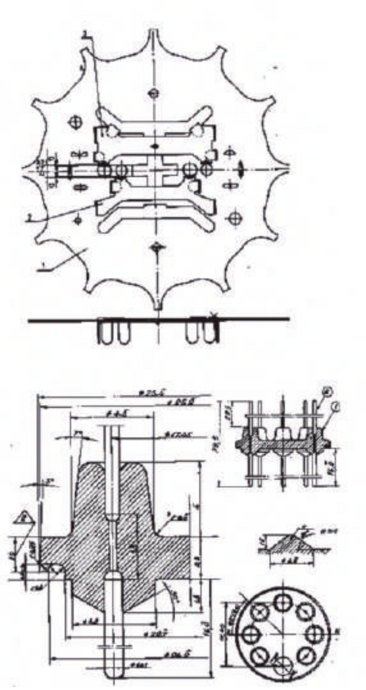


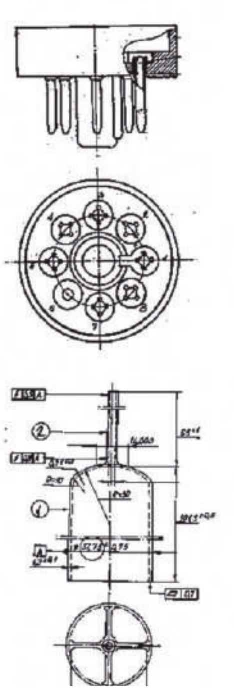
The heaters
(filaments) are made of Wo and Mo wire and coated with Al_2O_3 suspension by electrophoresis.

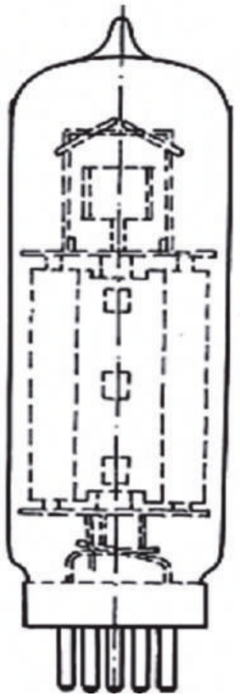


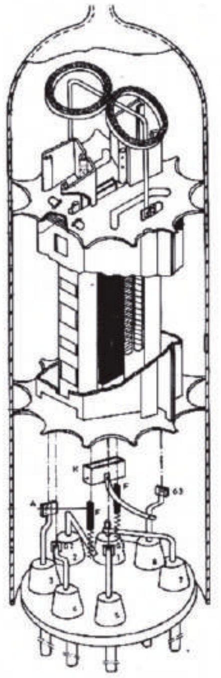
The grids are made of Mo wire and the supports are made of Cu Fe wire.

All the mentioned parts, as well as some others, are mounted on mechanically very precisely made mica plates (their shape, pattern and dimensions determine the geometry of the electrode system of the tube). The assembly made in this way is inserted into a previously made glass balloon, and further application of technological procedures such as vacuuming, sealing, getter burning, cathode activation, and quality control completes the production process of final product – a vacuum tube.









3.2 TRIODE CHARACTERISTICS

STATIC CHARACTERISTICS

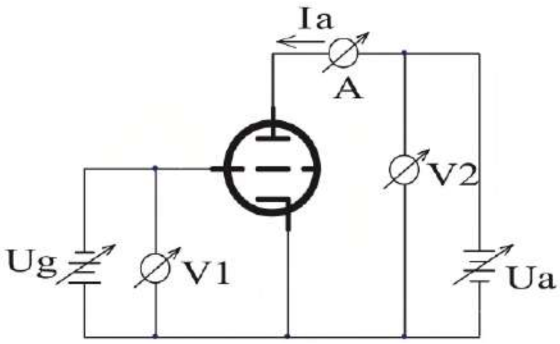
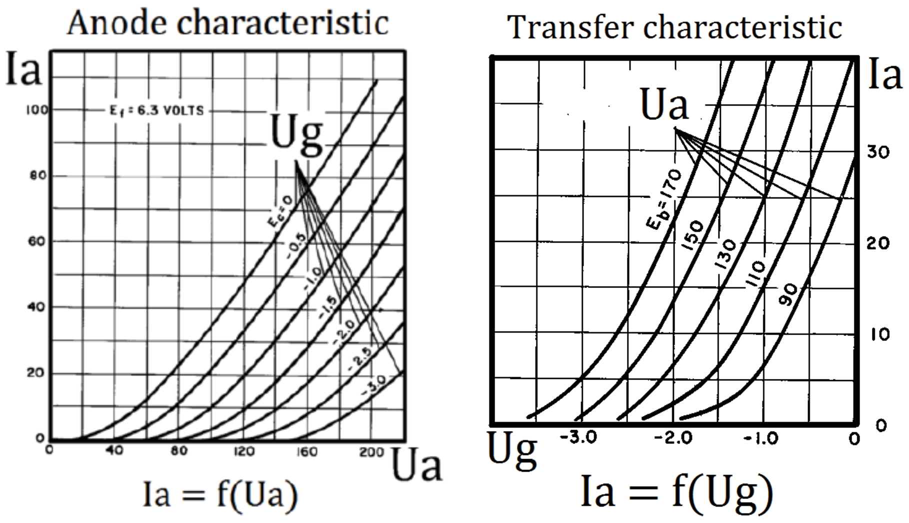


Fig. 3-02

The graph of the family of curves representing the anode DC current as a function of the set of DC voltages between the anode and cathode and the DC voltages between the grid and cathode is the so-called **triode static characteristics diagram**.

** In the text, the terms anode (grid) voltage are used as common expressions for the difference between the electrical potentials of the anode (grid) and cathode, in other words, the anode (grid) voltage is the electrical potential of the anode (grid) relative to the cathode.*



a. Characteristics **b. Characteristics**

Fig. 3-03

Graph of the anode current as a function of the anode voltage for several fixed different values of the grid voltage as parameters: **I_a= f (U_a)** |**U_g = const.** – so-called **static anode characteristics** of the triode (Fig.3-03a).

Graph of the anode current as a function of the grid voltage for several fixed different values of anode voltage as parameters: **I_a= f (U_g)** |**U_a = const.** - so-called **static transfer characteristics** of the triode (Fig.3-03b).

At a sufficiently high negative grid potential – grid voltage (**cut-off voltage**) the anode current is equal to zero.

The negative grid voltage necessary for the triode to operate in this limit state (cut-off) depends on the anode voltage.

When operating in the limit state (**cut-off)**, the triode is an open circuit.

At zero grid voltage (U_g = 0 V), the **conductivity** of the triode is **limited** by the **space charge** inside the tube. At this operating condition, the triode has maximum conductivity (minimum internal resistance).

The regular operation of the triode is limited by the cut-off voltage and the voltage limited by the space charge

(U_g = 0 V).

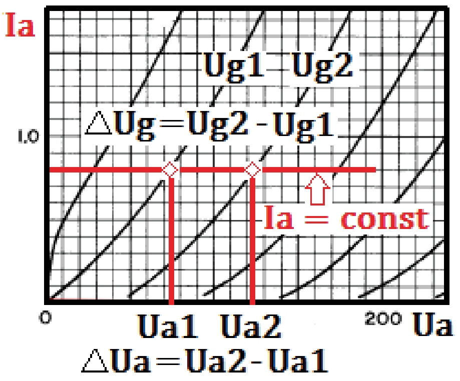
Anode characteristics and transfer characteristics can be used as very useful tools in the tube circuits design process.

Tube manufacturers publish the characteristics of the tube (**data sheet**) for each type of tube.

3.3 TRIODE PARAMETERS (μ, S, R_i)

μ - amplification coefficient (factor)

$$\mu = -\frac{dU_a}{dU_g}\,\Big|\,I_a = const.$$



Basic definition:
The amplification coefficient is the ratio of the anode potential to the negative potential of the grid that cause an electrical field at surface of the cathode equal to zero.

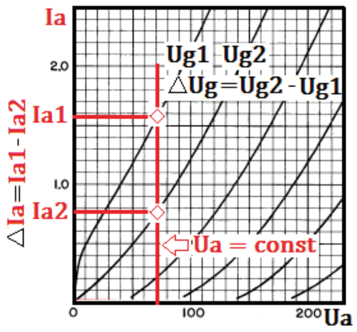
More suitable definitions (from a practical point of view):
The amplification coefficient of a triode is equal to the ratio of a change in anode voltage to the change in grid voltage required to cause the same change in anode current.

The amplification coefficient is equal to the ratio of the change in anode voltage (ΔU_a) to the change in negative grid voltage (ΔU_g), under the condition that the anode current does not change, i.e. it is constant

(I_a = const.).

In other words, the amplification coefficient represents the relative effect of the anode and grid voltage on anode current.

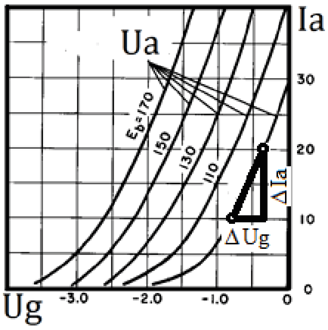
μ is the maximum theoretical voltage amplification of the tube.



S or g_m - Transconductance – mutual conductance (mA/V)

$$S = -\frac{dI_a}{dU_g}\,\Big|U_a = const.\,,\ g_m = \left(\frac{\Delta I_a}{\Delta U_a}\right)\ \Big|\,U_a = const.$$

The transconductance is defined as the ratio of a small change in the anode current to a corresponding small change in the grid voltage, under the condition that the anode voltage does not change, i.e. it is constant.

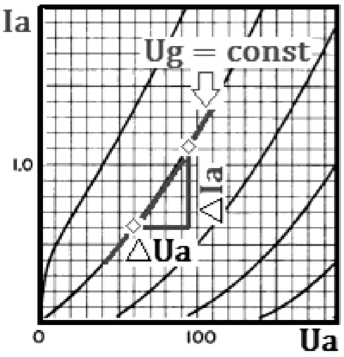


The transconductance (from a practical point of view) is equal to the change in the anode current (mA) caused by a change in the grid voltage of 1V, under the condition that the anode voltage does not change, i.e. it is constant. Transconductance represents the variation of the anode current caused by the variation of the grid voltage, under the condition that the anode voltage does not change, i.e. it is constant.

The unit for transconductance is the mho (Siemens [S]), μmho = 10^{-6} S (commonly: mA / V or μmho *).*

The value of the transconductance is equal to the slope of transfer characteristic of the triode.

R_i - Internal resistance (plate or anode resistance) (Ω)



$$R_i = -\frac{dU_a}{dI_a}\,\Big|U_g = const.$$

The internal resistance is defined as the ratio of a small change in the anode voltage to the resulting change in the anode current, under the condition that the grid voltage does not change, i.e. it is constant.

*The value of the **R_i** is equal to the reciprocal of the gradient (slope) of the anode characteristic curve (1 / (ΔI_a / ΔU_a) = ΔU_a / ΔI_a).*

The parameter R_i differs from, conditionally speaking, DC resistance of the triode (the ratio of DC anode voltage to DC anode current; static operation of the triode). It is more exact to describe R_i as the dynamic internal resistance of the triode (anode resistance).

The R_i of a triode is the resistance that opposes the AC current flowing through it.

Relations between parameters

$$\mu = \boldsymbol{S}\left[\frac{mA}{V}\right] \times \boldsymbol{R_i}[\boldsymbol{k}\Omega] \;; \qquad \mu = \boldsymbol{g_m}[\mu\Omega] \times \mathbf{10^{-6}} \times \boldsymbol{R_i}[\Omega]$$

Parallel connection (two or more tubes)

$R_{i\,eq}$ – Equivalent internal resistance:

$$\frac{1}{R_{i\,eq}} = \frac{1}{R_{i1}} + \frac{1}{R_{i2}} + \cdots + \frac{1}{R_{in}} = \sum_{i=1}^{n} \frac{1}{R_i} \;; \tag{301}$$

R_i –internal resistance of each tube connected in parallel

S_{eq} – Equivalent transconductance

$$S_{eq} = S_1 + S_2 + \cdots + S_n = \sum_{i=1}^{n} S_i \;; \tag{302}$$

S_i – transconductance of each tube connected in parallel

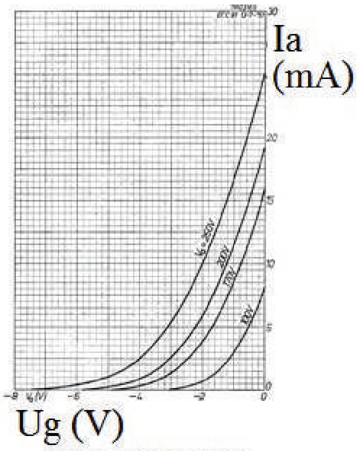
μ_{eq} – Equivalent amplification coefficient (two tubes connected in parallel)

$$\mu_{eq} = \frac{\mu_1 \times R_{i2} + \mu_2 \times R_{i2}}{R_{i1} + R_{i2}} \tag{303}$$

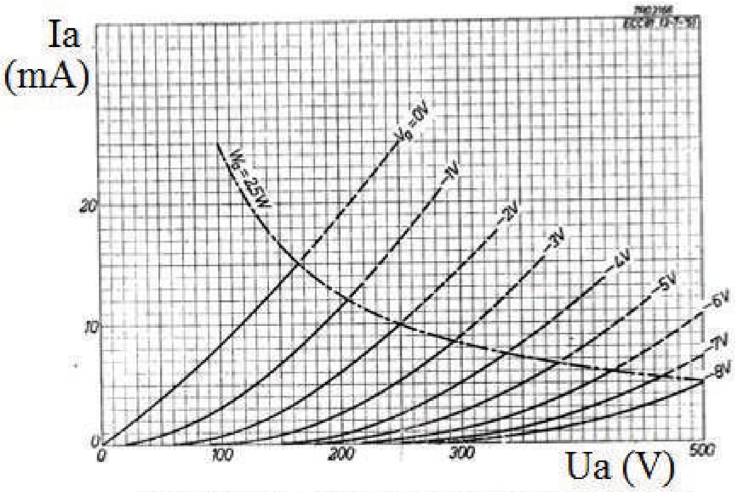
Two tubes with identical characteristics connected in parallel:

$$R_{i\,eq} = \frac{R_i}{2} \qquad\qquad S_{eq} = 2 \times S \qquad\qquad \mu_{eq} = \mu$$

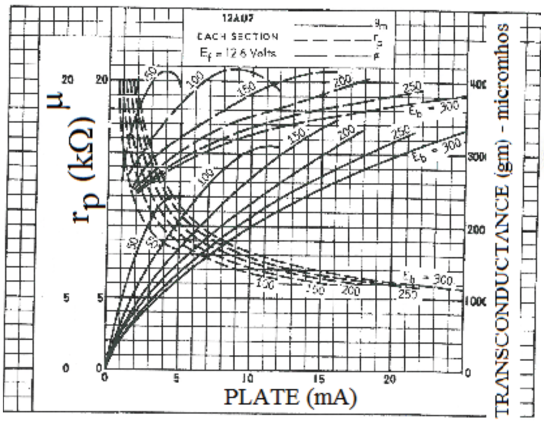
Typical diagrams published in tube data sheets:



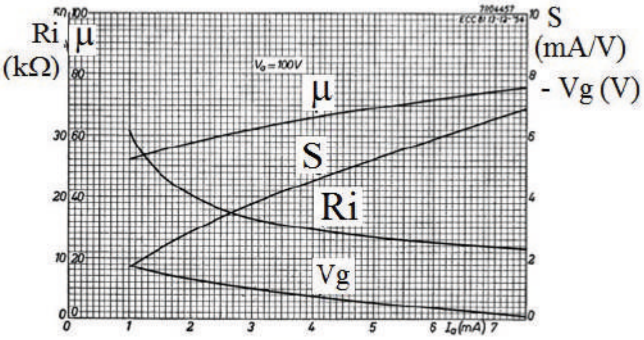
TRANSFER CHARACTERISTIC



ANODE CHARACTERISTIC



μ, g_m, r_p



μ, Ri, S

3.4 AMPLIFYING EFFECT OF THE TRIODE

*** analysis for small AC signals**

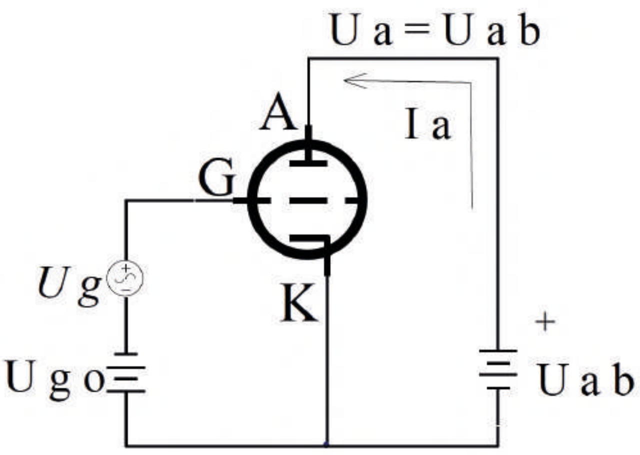


Fig. 3-04

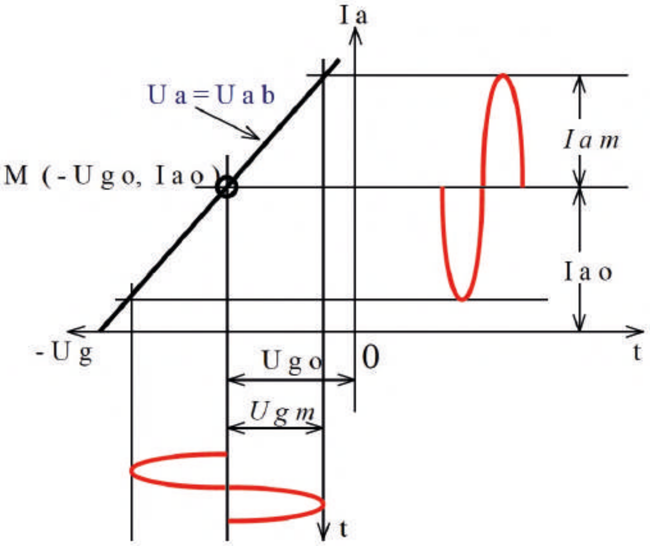


Fig. 3-05

- The anode current is determined by the anode voltage and the grid voltage Ia = f (Ua, Ug) and if the value of any of them changes, the value of the anode current also changes.

- In the static mode of operation of the triode, the anode DC current (**Iao**) is determined by the anode constant DC voltage (**Ua = Uab)** and the grid negative constant DC voltage at (- **Ugo**).
  Point **M** on the graph of the transfer characteristic (**Ia = f (Ug**). **Ua = Uab**, Fig. 3-05), determined by the grid DC voltage **Ugo** and the anode DC current **Iao**, i.e. **M (Ugo, Iao**) is the so-called **working** (operating) point of the tube.

- At a constant value of the anode voltage (power supply voltage), any change in the value of the grid voltage causes a change in the value of the anode current (**Ia = f (Ug**), **Ua = Uab = constant**).

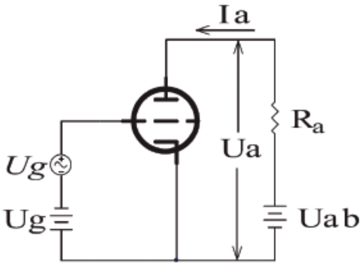
- If the AC voltage source (the ″signal″ to be amplified), ***Ug = Ugm • cos(ωt)*** is connected in series with the DC voltage source in the grid circuit, the anode current changes depending on how the grid voltage changes, i.e.:

$$u_g = -U_{g0} + U_{gm} \times \cos(\omega t) \tag{304}$$

$$i_a = -I_{a0} + I_{am} \times \cos(\omega t) \tag{305}$$

**** Fig. 3-05: a small change in grid AC voltage causes a large change (of the same shape as the change in grid voltage) in the anode current.***

3.4.1 VOLTAGE AMPLIFICATION



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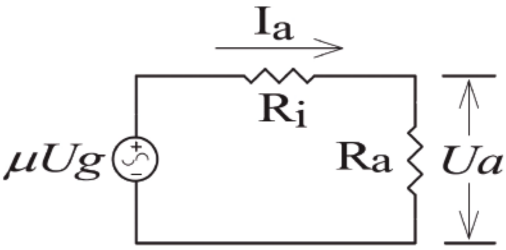


Fig. 3-06

Fig. 3-07

- A vacuum **tube** can be used as a **voltage amplifier** by placing a resistance $\mathbf{R_a}$ between the anode power supply (U_{ab}) and the anode of the tube, Fig. 3-06.

- The resistance $\mathbf{R_a}$ changes the operating conditions of the tube: the anode DC voltage ($\mathbf{U_a}$) is not equal to the DC voltage of the anode power supply ($\mathbf{U_{ab}}$), i.e. U_a is lower than the voltage $\mathbf{U_{ab}}$ by the value of the voltage drop across the resistance $\mathbf{R_a}$ caused by the current (I_a) flowing through the tube:

$$U_a = U_{ab} - R_a \times I_a \tag{306}$$

- If the grid voltage changes, the anode voltage does not remain constant, it also changes depending on the change in the voltage drop across the resistance $\mathbf{R_a}$ in the anode circuit caused by the change in the current flowing through the tube.

- The working (operating) point does not move along one static characteristic, it moves over several static characteristics (from one to the other) that correspond to the instantaneous values of the anode voltage.

Anode current is:

$$i_a = S \times u_g + \frac{u_a}{R_i} \tag{307}$$

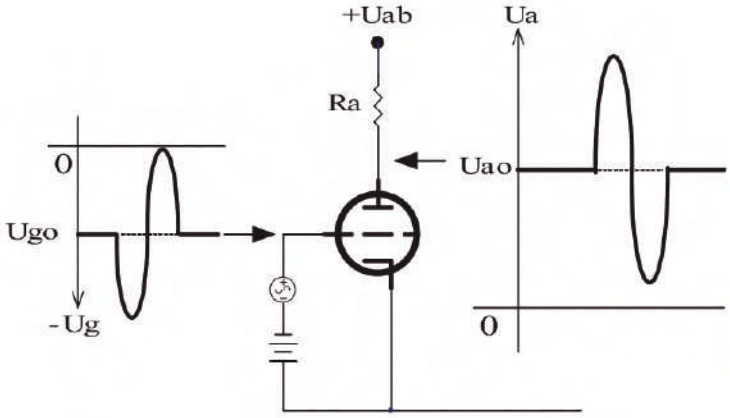
The AC voltage across the resistance $\mathbf{R_a}$ is:

$$u_a = -(R_a \times i_a) \tag{308}$$

The " **-** " sign means a phase change, i.e. the input signal (voltage) $\boldsymbol{u_g}$ at the grid and the output signal (voltage) $\boldsymbol{u_a}$ are **opposite in phase**.

Voltage amplification (A) is:

$$A = \frac{u_a}{u_g} = \frac{-(S \times u_g + \frac{u_a}{R_i}) \times R_a}{u_g} = -\frac{S \times R_i \times R_a}{R_i + R_a} = -\mu \times \frac{R_a}{R_i + R_a} \tag{309}$$



*# Fig. 3-07 represents the so-called **equivalent circuit** where the tube is substituted by a voltage generator –* μUg *connected in series with the internal resistance of the triode and all other impedances of the circuit.*

The *equivalent circuit can be used to analyze AC signal amplifier circuits. DC sources are omitted (they are short circuits for the AC signals).*

The analyze of an AC signal amplifier circuit using the equivalent circuit is much simpler because the tube amplifier circuit can be treated as a passive electric circuit and the basic electric laws can be applied: Ohm's law, Kirchhoff's circuit laws, Thevenin's theorem…

*# The result of a more detailed analysis that considers the **internal capacitances** (the capacitances between the electrodes of the tube) of the tube and the impedance **Z_a** of the anode circuit is that the amplification is a complex number.*

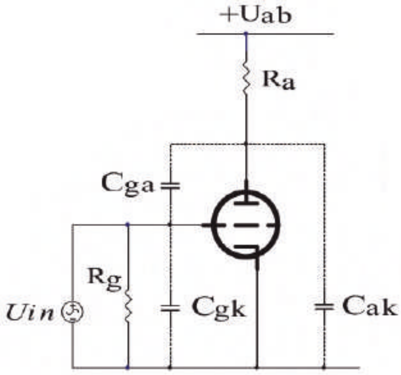


Fig. 3-08

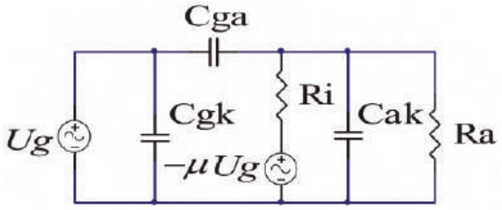


Fig. 3-09

Complex number: ***z*** *= a + jb****;*** ***a –*** *real part;* ***b –*** *imaginary part;* ***j –*** *imaginary unit****;*** ***j^2 = - 1***

Complex number – trigonometric form:

$$z = a + jb = r \times (\cos\varphi + j\sin\varphi),\ r = +\sqrt{a^2+b^2};\ \tan\varphi = \frac{b}{a};$$

Anode circuit Impedance:

$$\frac{1}{Z_a} = \frac{1}{R_a} + \frac{1}{jX_a}$$

Amplification:

$$A = A_r + jA_j$$

Real part of the amplification:

$$A_r = -\frac{\mu\times(\frac{R_i}{R_a}+1)}{(\frac{R_i}{R_a}+1)^2+(\frac{R_i}{X_a})^2}$$

Imaginary part of the amplification:

$$A_j = -\frac{\mu\times\frac{R_i}{R_a}}{(\frac{R_i}{R_a}+1)^2+(\frac{R_i}{X_a})^2}$$

Input capacitance:

$$C_{in} = C_{gk} + C_{ga} \times [1 + \frac{\mu\times\left(\frac{R_i}{R_a}+1\right)}{\left(\frac{R_i}{R_a}+1\right)^2+\left(\frac{R_i}{X_a}\right)^2}]$$

The increase in input capacitance of a tube amplifier circuit is known as the Miller effect.

The maximum possible value of the input capacitance is: $C_{in} = C_{gk} + C_{ga} \times (\mu + 1)$

Input capacitance of the amplifier circuit is: $\boldsymbol{C_{in} = C_{gk} + C_{ga} \times (A + 1)}$

** A – amplification of the amplifier circuit*

It is very important to pay attention to the Miller effect in the process of designing of multi-stage amplifiers, due to the great effects of Miller capacitances on the amplitude characteristics of the amplifier, especially at high frequencies.

3.5 WORKING POINT (QUIESCENT POINT) AND WORKING (LOAD) LINE

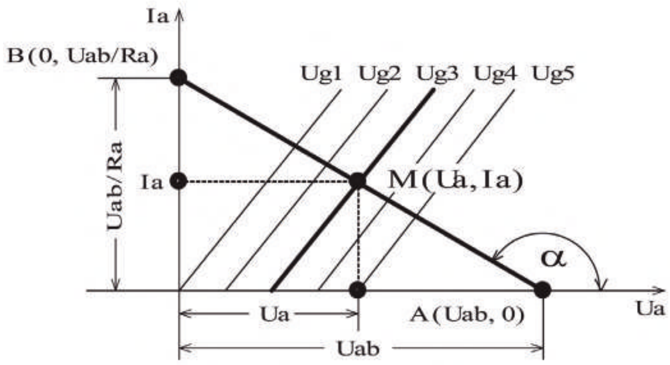


Fig. 3-10

The anode DC voltage is determined by the equation:

$$U_a = U_{ab} - R_a \times I_a \tag{310}$$

To determine $\boldsymbol{U_a}$ and $\boldsymbol{I_a}$ it is necessary to know the static characteristics of the anode:

$\mathbf{I_a = f\ (U_a)}$, Fig. 3-10.

The working point is located at the intersection of the characteristic $\mathbf{I_a = f\ (U_a)}$ [$\mathbf{U_g = constant}$] and the line defined by equation (310):

$$I_a = \frac{U_{ab}}{R_a}$$

Drawing the working line (load line):

- Draw point **A** on the abscissas**:** value equal to $\mathbf{U_{ab}}$ (DC voltage of the anode power supply), [**A** (U_{ab}, 0)]
- Draw point **B** on the ordinate**:** value equal to $\mathbf{U_{ab}\ /\ R_a}$, [**B** (0, U_{ab} / R_a)]
- Draw a straight **line (working line)** through the points **A** and **B**.

The straight line determines by the points A (U_{ab}) and B (U_{ab} / R_a) through which it is drawn is the **working (load) line** and can be expressed mathematically in the form of a function:

$$I_a = \frac{U_{ab}-U_a}{R_a} \tag{311}$$

The **working point M ($\mathbf{U_a}$, $\mathbf{I_a}$)** [**quiescent point**] is located at the intersection of the working line and the static anode characteristic corresponding to the chosen negative grid voltage (U_g). The coordinates of the point M determine the DC anode voltage $\mathbf{U_a}$ and the DC anode current $\mathbf{I_a}$ of the triode at the working point.

Drawing the dynamic characteristic (dynamic working line) on the diagram $I_a = f(U_g)$

Transfer all intersection points of the line **U_a = const**. (a straight line drawn from the point **U_a** parallel to the **I_a** axis) with static characteristics of the anode characteristics diagram **Ia = f (Ua)** to the transfer characteristics diagram **Ia = f (Ug)**.

Draw the **static characteristic** through the transferred points on the transfer characteristics diagram **$I_a = f(U_g)$**.

Transfer all intersection points of the working line with static characteristics of the anode characteristics diagram **$I_a = f(U_a)$** to the transfer characteristics diagram **$I_a = f(U_g)$**.

Draw the **dynamic characteristic** (bold line) through the transferred points on the **$I_a = f(U_g)$** diagram.

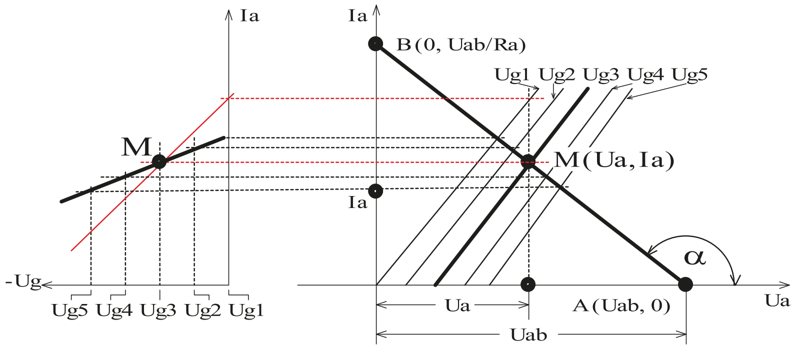


Fig. 3-11

*The slope of the **static transfer** (mutual) characteristic is:*

μ / R_i .

*The slope of the **dynamic** characteristic is:*

$\mu / (R_i + R_a)$

 # The slope of the dynamic characteristic of the triode is lower than the slope of the static characteristic.

3.6 COMPLETE AMPLIFICATION MECHANISM

The signal at the anode is of the same shape but opposite in phase and of a significantly higher amplitude compared to the AC signal applied to the triode grid (amplification: A = U_{am} / U_{gm}).

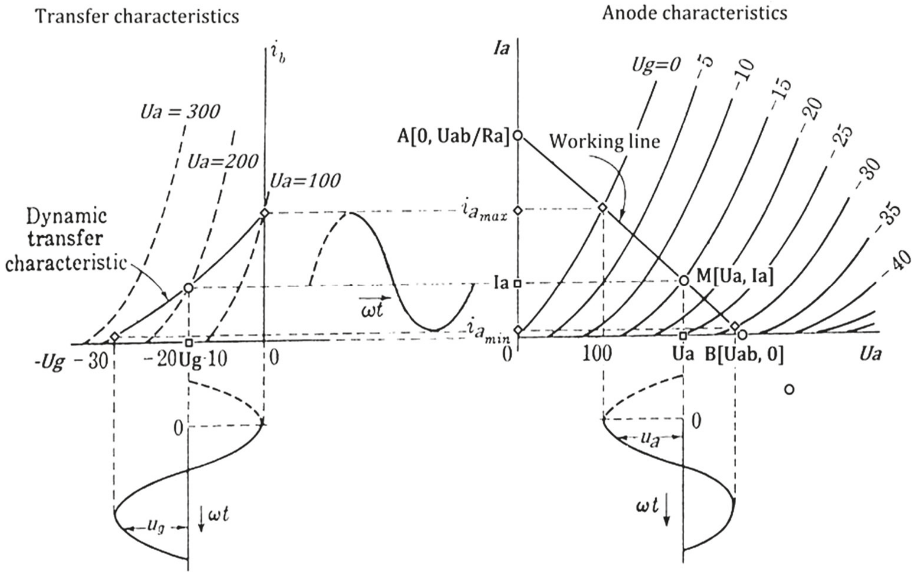


Fig. 3-12

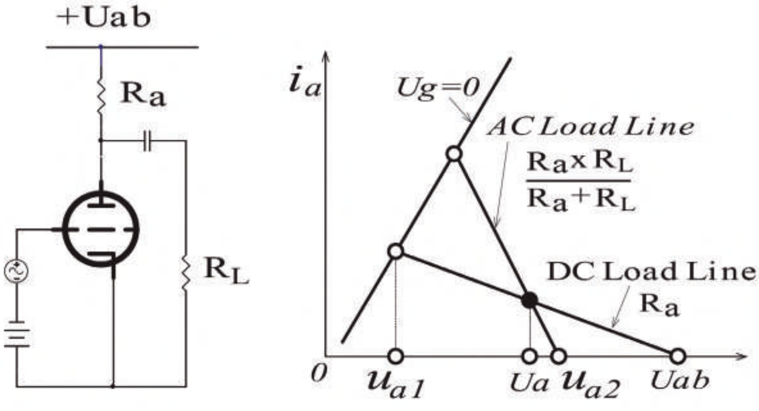
 # Note: The technical characteristics of the triode amplifier circuit depend on the correct determination of the triode operating point (choice of the adequate value of the resistor Ra and correct setting of the electrical potential of the grid of the triode – grid bias voltage).

In practice, the signal amplification effect of the triode circuit is used to amplify the input signal and to supply some load R_L with the amplified signal.

- -

Determination of the anode load R_a for maximum output voltage without exceeding a given distortion.

(Richard Q Twiss)



The analysis is based on the presumption that the positive and negative swings of the anode voltage, measured from the quiescent point, are equal and limited by the cut-off voltage and the zero grid voltage (saturation), respectively.

Fig. 3-13

U_{ab} – power supply DC voltage; U_a – anode DC voltage; I_a – anode DC current; R_L – Load resistance; $R'_a = \frac{R_a \times R_L}{R_a + R_L}$;

u_a – anode instantaneous voltage; i_a – anode instantaneous current;
u_{a1} – anode voltage at $U_g = 0$; u_{a2} – anode voltage at cut-off;

Anode instantaneous voltage at cut-off is: $u_{a2} = U_a + I_a \times R'_a$

Anode instantaneous current at $U_g = 0$ is: $i_a = \frac{u_{a1}}{R_i}$

The equation of the AC load line is: $u_a - U_a = -(i_a - I_a) \times (R'_a)$

By solving the above system of equations for u_{a1}: $$u_{a1} = \frac{U_a + I_a \times R'_a}{1 + \frac{R'_a}{R_i}} \quad (Eq.I)$$

Positive and negative swings of the voltage U_a must be equal: $u_{a2} = U_a - I_a \times R'_a \quad (Eq.II)$

By eliminating u_{a1} in (Eq.I) and (Eq.II) and combining with the equation of the DC working line: $\boldsymbol{U_a = U_{ab} - I_a \ x \ R_a}$, it is possible to determine I_a and calculate the peak output voltage: $u_{a\,peak} = I_a \, x \, R'_a$.

The effective output voltage is: $u_{a\,RMS} = \frac{u_{a\,peak}}{\sqrt{2}}$ *i.e.:* $$u_{a\,RMS} = \frac{\frac{U_{ab}}{\sqrt{2}}}{1 + \frac{R_a}{R'_a} + \frac{2 \times R_i}{R'_a}} = \frac{\frac{U_{ab}}{2 \times \sqrt{2}}}{1 + \frac{R_i}{R_L} + \frac{R_i}{R_a} + \frac{R_a}{2 \times R_L}}$$

Maximizing the output voltage ($u_{a\,RMS}$) with respect to R_a :

$$R_a = \sqrt{2 \times R_i \times R_L}$$

and

$$u_{RMS\,max} = \frac{\frac{U_{ab}}{2 \times \sqrt{2}}}{1 + \frac{R_i}{R_L} + \sqrt{\frac{2 \times R_i}{R_L}}}$$

The derivation may be generalized by considering nonlinearity of the tube characteristics. If the negative swing of the anode voltage is a *times greater than the positive swing:*

$$R_a = \sqrt{(1+a) \times R_i \times R_L} \quad \text{and} \quad u_{RMS\,max} = \frac{\frac{U_{ab}}{2\times\sqrt{2}}}{1+\frac{R_i}{R_L}+2\times\sqrt{\frac{R_i}{(1+a)\times R_L}}}$$

- -

3.7 BIAS

In a practical triode circuit, the negative electrical potential of the grid relative to the potential of the cathode can be realized by using a separate DC voltage source connected in the grid circuit, the so-called **fixed bias**, or by using a resistor R_k connected in the cathode circuit (between the cathode and the ground or reference point), the so-called **automatic bias** (more convenient in practice).

 # Automatic bias:

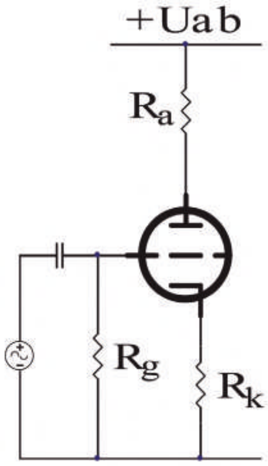
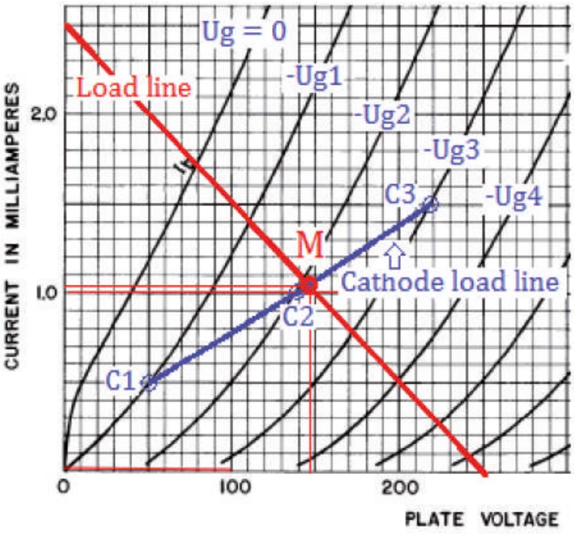


Fig. 3-14

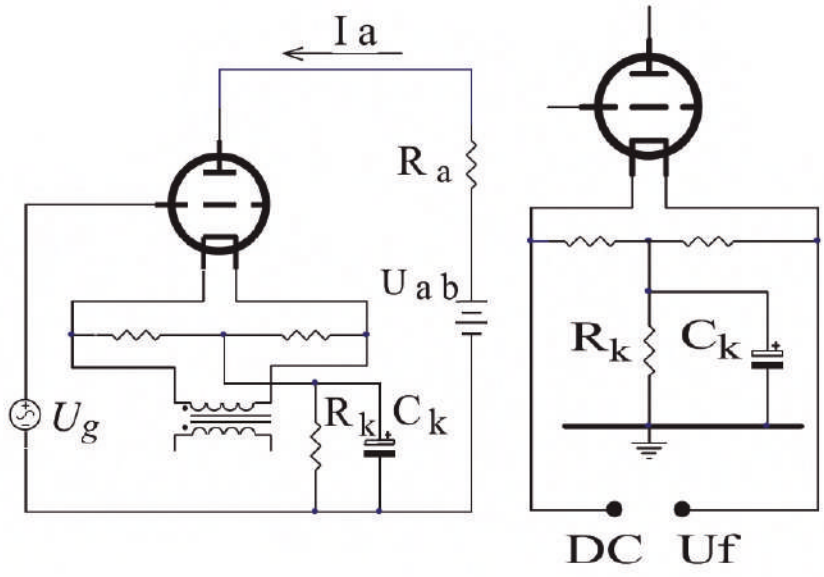
The electrical current flowing through the tube causes a voltage drop across the cathode resistor and the electrical potential of the cathode becomes positive relative to the ground – reference potential. As the grid resistor (connected between the grid and the ground point) keeps the grid at ground potential, the grid potential becomes negative relative to the cathode.

$$-U_{g\text{-}k} = R_k \times I_{a\,(DC)}$$



R_k is determined by the values **I_a** and **U_g** at the quiescent point **M**: $R_k = \frac{-U_g}{I_a}$ **.** Graphical analysis (diagram above) starts by drawing a load line. Since the grid bias voltage depends on the anode current, it is necessary to draw another line, the so-called **cathode load line.** The cathode load line is drawn by calculating and drawing the points $I_a = \frac{-U_g}{R_k}$ i.e.: C1 (U_{g1} / R_k) and C3 (U_{g3} / R_k), [$U_{g1} < U_g < U_{g3}$] and connecting them with a straight line. The intersection of the cathode load line with the load line is the **quiescent point (M).**

Automatic bias (directly heated cathode)



Note:

- The grid voltage contains an AC voltage component caused by the AC power supply to the heater.
- The amplified signal contains an AC component caused by the AC power supply to the heater.

Using the heater AC supply circuit configuration (directly heated cathode circuit) Fig. 3-15 **suppresses the fundamental frequency and odd harmonics of the supply voltage.**

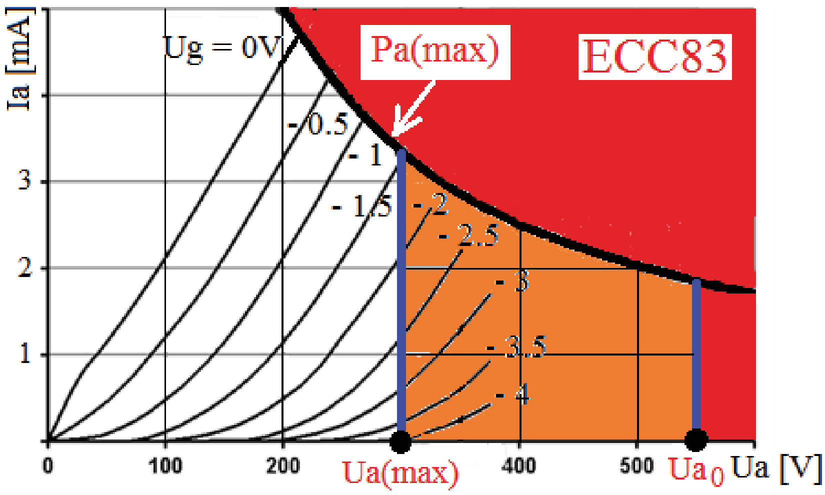
The **even** harmonics are not suppressed. The anode voltage contains the **even** harmonics of the AC voltage.

Fig. 3-15

A *DC power supply to the heater can also be used.*

3.8 PRECAUTIONS WHEN DETERMINING THE OPERATING CONDITIONS OF THE TUBE
(THE QUIESCENT POINT - BIAS AND THE LOAD LINE)

The load line and the quiescent point must be placed below the power dissipation curve ($P_{a(max)}$) on the diagram of the anode characteristics $I_a = f(U_a)$.



The anode voltage must never exceed U_{a0} (max. anode voltage).

The anode voltage at the quiescent point must not exceed $U_{a(max)}$

(max. permissible anode voltage at the quiescent point, i.e. the maximum DC voltage at which the tube can operate continuously).

Fig. 3-16

Note: Maximal anode dissipation $P_{a(max)}$ and max. anode voltage V_{a0} $V_{a(max)}$ (maximum ratings) as well as other characteristic tube parameters under typical operating conditions (typical operation) are usually published in the data sheets of the tube manufacturers.

3.9 AMPLIFIER CLASSES OF OPERATION

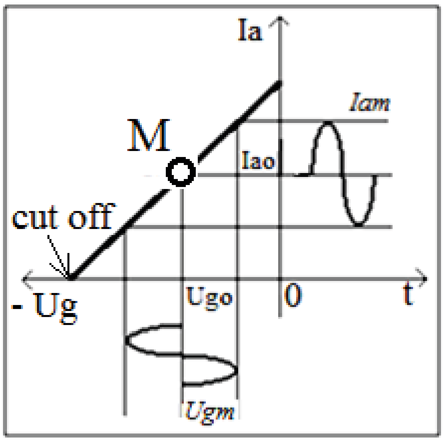


Fig. 3-17

- **Class A:**

The working point (quiescent point) is located somewhere around the middle of the transfer characteristic $I_a = f(U_g)$, *(the middle of the most linear part of the transfer characteristic).*

The signal voltage applied to the grid always fulfills the condition:

$U_{gm} < U_{g0}$, i.e. the tube does not operate either in the saturation region or in the region close to the cut-off voltage.

- Low nonlinear distortion.
- Low output power.
- ***Low efficiency.***

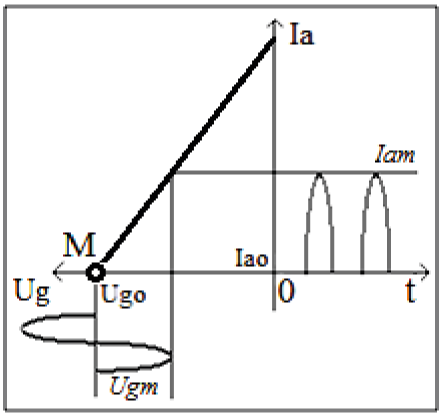


Fig. 3-18

- **Class B:**

The working point is located at the knee of the characteristic $I_a = f(U_g)$ i.e. $U_{g0} = U_{g\ cut\text{-}off}$.

Anode DC current at working (quiescent) point (signal AC voltage is not applied to the grid) is approximately zero. **M ($-U_{g(cut\ off)}$, $I_a = 0$)**.

Anode current flows only when an AC signal is applied to the grid for a period of time approximately equal to half of each cycle of the applied AC voltage.

Efficiency and output power are higher than Class A.

In audio amplifiers, Class B can be used in a Push-Pull amplifier configuration only.

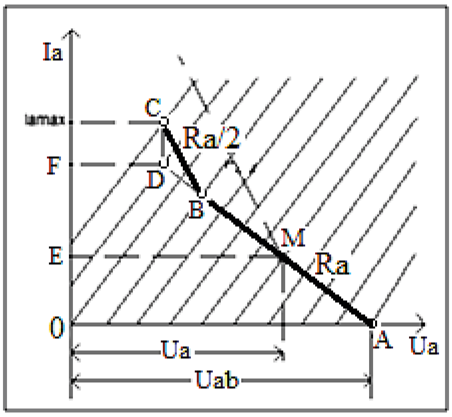


Fig. 3-19

- **Class AB:**

The working point is located somewhere between the Class A and Class B working (quiescent) points in on the transfer characteristic.

The amplifier operates in Class A at low AC voltages of the input (driving) signal and in Class B at high levels of the input signal.

In a push pull amplifier configuration at low driving voltages, the working point of one tube moves along the part **MA** of the load line, and at the same time the working point of the other tube moves along the part **MB** of the load line defined by R_a on the anode characteristic diagram.

The anode current of one tube decreases as the negative amplitude of the driving voltage increases and finally stops flowing at point A on the anode characteristic diagram *(after that only half of the primary winding of the output transformer is used and other tube “sees” a different impedance at its anode).*

To the left of the point B, the anode load changes from R_a to a value equal to $R_a / 2$.(**In reality, the load line is not a straight line, it is curved).**

3.10 AMPLIFIER DISTORTION TYPES

In general - distortion can be defined as a measure of the deviation of the signal waveform at the amplifier output compared to the signal waveform at the amplifier input.

Non-linear distortion – caused by non-linearity of the tube characteristics $I_a = f(U_a)$, $I_a = f(U_g)$ which are not straight lines, so the tube parameters (μ, S, R_i) are not constant along the characteristics.

They can be:

- Harmonic
- Non-harmonic

Linear or amplitude_distortion – caused by the characteristics of the electronics components used in the amplifier circuits, the layout of the components, the wiring and the internal capacitances of the tubes.

3.10.1 HARMONIC DISTORTION

If the input signal is a periodic function in time, the amplified signal at the anode circuit contains the fundamental harmonic and higher order harmonics.

If the characteristic $I_a = f(U_g)$ is approximated by a third-order polynomial:

$$i_a = I_{a0} + a_1 u_g + a_2 u_g^2 + a_3 u_g^3$$

and the input signal is: $u_g = U_{gm} \cos(\omega t)$

After mathematical derivations, the function of the anode current can be expressed in the form:

$$i_a = I_{a0} + \frac{1}{2} a_2 U_{gm}^2 + \left(a_1 U_{gm} + \frac{3}{4} a_3 U_{gm}^3\right) \cos \omega t + \frac{1}{2} a_2 U_{gm}^2 \cos 2\omega t + \frac{1}{4} a_1 U_{gm}^3 \cos 3\omega t$$

The anode current is the sum of the:

- DC component: $I'_{a0} = I_{a0} + \frac{1}{2} a_2 U_{gm}^2$
- First harmonic: $I_{a1} = \left(a_1 U_{gm} + \frac{3}{4} a_3 U_{gm}^3\right) \cos \omega t$
- Second harmonic: $I_{a2} = \frac{1}{2} a_2 U_{gm}^2 \cos 2\omega t$
- Third harmonic: $I_{a3} = \frac{1}{4} a_3 U_{gm}^3 \cos 3\omega t$

* *The DC component of the anode current also changes – this means that the working point changes its position. Using a fixed grid bias causes a larger change in anode DC current than using an automatic grid bias.*

In order to quantify the effect of tube characteristic nonlinearity on signal distortion, a measure commonly used in real measurements is defined – **Distortion Coefficient** or **Clear Factor.**

The Distortion Coefficient or Clear Factor of a harmonic is defined as the ratio of the effective value of the harmonic of the anode current and the total effective value of anode current:

$$k_n = \frac{I_{n\omega}}{\sqrt{I_{\omega}^2 + I_{2\omega}^2 + I_{3\omega}^2 + \cdots}} \times 100\%$$

Total Clear Factor: the ratio of the sum of the effective values of higher-order harmonics and the effective value of the total anode AC current (in percent):

$$k = \sqrt{k_2^2 + k_3^2 + \cdots + k_n^2}$$

3.10.2 HARMONIC DISTORTION — GRAPHICAL ANALYSIS

Graphical analysis is based on the approximation of characteristic ***$I_a = f\,(U_g)$*** by a second-order polynomial:

$$i_a = I_a + A_0 + A_1 \cos \omega t + A_2 \cos 2\omega t$$

Fig. 3-12:

For: **$\omega t = 0$** → $i_a = i_{amax}$
For: $\omega t = \pi / 2$ → $i_a = I_a$
For: **$\omega t = \pi$** → $i_a = i_{amin}$

and:

$$i_{a\,max} = I_a + A_0 + A_1 + A_2$$

$$I_a = I_a + A_0 + A_2 \quad \rightarrow \quad A_0 = A_2$$

$$i_{a\,min} = I_a + A_0 - A_1 + A_2$$

Second Harmonic Distortion (Clear Factor) - (HD_2):

$$HD_2 = k_2 = \frac{A_2}{A_1} 100\% = \frac{1}{2} \times \frac{i_{a\,max} + i_{a\,min} - 2I_a}{i_{a\,max} - i_{a\,min}} \times 100\%$$

Fig. 3-20:

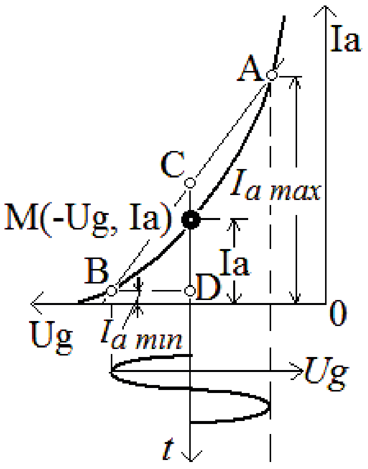


Fig. 3-20

$$\overline{CD} = \frac{i_{a\,max} - i_{a\,min}}{2} = A_1$$

$$\overline{CM} = \frac{1}{2}(i_{a\,max} - 2I_a + i_{a\,min}) = 2A_2$$

Second Harmonic Distortion (Clear Factor):

$$\boldsymbol{HD_2 = k_2 = \frac{A_2}{A_1} = \frac{CM}{2CD} \times 100\%}$$

3.10.3 NON-HARMONIC DISTORTION

Non-linear non-harmonic distortions: the anode current (output signal) contains not only the frequencies that are multiples of the fundamental frequency, but also contains: the fundamental harmonic and harmonics of a higher order, sums and differences of the fundamental harmonic and harmonics of a higher order, caused by the non-linear characteristic of the tube.

Coefficient of non-linear non-harmonic distortion, M:

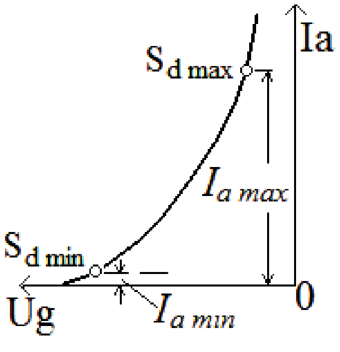


Fig. 3-21

$$M = \frac{S_{d\,max} - S_{d\,min}}{S_{d\,max} + S_{d\,min}} \times 100\%$$

S_{dmax} – max. dynamic transconductance corresponding to the max. anode current:

$$S_{d\,max} = \frac{I_{a\,max}}{U_{gm}}$$

S_{dmin} – min. dynamic transconductance corresponding to the min. anode current:

$$S_{d\,min} = \frac{I_{a\,min}}{U_{gm}}$$

Example:

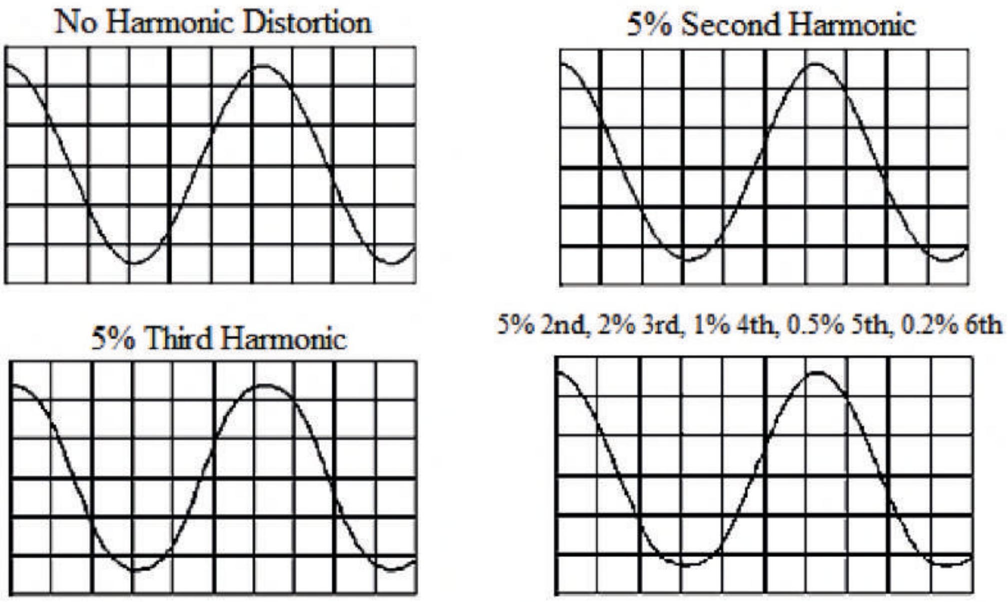


Fig. 3-22

3.10.4 LINEAR (AMPLITUDE; FREQUENCY) DISTORTION

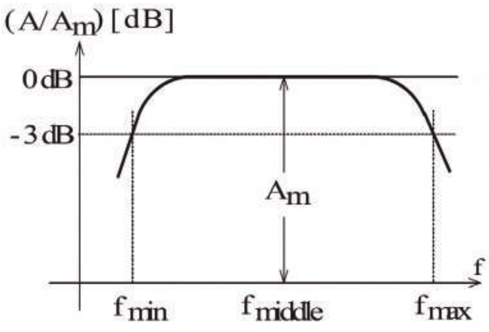


Fig. 3-23

Linear distortion occurs when the amplification of input signals of equal amplitudes and different frequencies is not equal over the entire defined frequency range, A (f) = A (f_a ... f_b)) ≠ const.

A measure of amplifier quality related to amplitude distortion is the so-called **amplitude characteristic** or **frequency characteristic** (or frequency response) of the amplifier.

The amplitude characteristic of an amplifier can be defined as the amplifier's amplification as a function of the frequency of the input signal.

Frequencies at which the amplification (A) is **3dB** lower than its nominal value (A_m - amplification at the middle frequencies) are the so-called **cut-off frequencies**:

f_{min} – low frequency cut-off, $(A / A_m) = - 3$ dB
f_{max} – high frequency cut-off, $(A / A_m) = - 3$ dB

3.11 European Tube Designation Code

The tube type designation consists of**: Two or more LETTERS followed by a serial NUMBER.**

L1	**L2 L3 …**	**N1 N2…**

L1

<u>First letter</u> indicates
HEATER
Voltage or Current

L2, L3, …

<u>Second and subsequent letters</u> indicate the construction and/or application of the tube (If there is more than one serial number, it indicates that there is more than one electrode system and these letters are placed in alphabetic order. Each letter indicates the corresponding electrode system)

N1, N2,…

<u>First number (N1)</u>
indicates
the type of the **BASE**

A	4 Volts
D	1.4 Volts
E	6.3 Volts
G	miscellaneous
L	450 mA
P	300 mA
U	100 mA

A	Diode
B	Double Diode
C	Triode
D	Power output triode
E	Tetrode
F	Pentode
H	Hexode or Heptode
K	Octode or Heptode
L	Power Output Tetrode or Power Output Pentode
M	Tuning indicator
Y	Half – Wave Rectifier
Z	Full – Wave Rectifier

1	miscellaneous
2	miniature 10 - pin
3	octal
5	magnoval
8	noval
9	miniature 7 – pin

SYMBOLS

Electrodes

a	ANODE
f	HEATER or FILAMENT
f_c	HEATER or FILAMENT TAP
g	GRID
i.c.	INTERNAL CONECTION
k	CATHODE
n.c.	TUBE PIN WHICH CAN BE CONNECTED EXTERNALLY
s	INTERNAL SHIELD

Miscellaneous

d_{tot}	Total distortion
f	Frequency
F	Noise factor
g	Voltage gain per stage
G	Power amplification (Gain)
S	Mutual conductance
S_c	Conversion conductance
η	Efficiency
μ	Gain or amplification factor
μ_{g2g1}	Gain or amplification factor of grid No. 2 with respect to grid No. 1

Voltages

V_a	Anode voltage
V_b	Supply voltage
V_{ba}	Anode supply voltage
V_{ap}	Peak value of anode voltage
V_{arms}	A.C. anode voltage
V_{eff}	RMS voltage value
V_f	Heater voltage
V_g	Grid voltage
V_{gp}	Peak grid voltage
V_{kf}	Voltage between cathode and heater (filament)
V_{kfp}	Peak value of voltage between cathode and heater (filament)
V_0	A.C. output voltage; D.C. output voltage (rectify)

Currents

I_a	Anode current
I_b	Supply current
I_{amax}	Anode current at full drive
I_{amin}	Anode current without drive
I_{ap}	Peak anode current
I_f	Heater or Filament current
I_g	Grid current
I_{gmax}	Grid current at full drive
I_{gmin}	Grid current without drive
I_{gp}	Peak grid current
I_k	Cathode current
I_0	D.C. current supplied by a rectifying tube

Capacitances

C_a	Anode to all other elements (electrodes) except control grid
C_{ag}	Anode to grid, all other elements (electrodes) earthed
C_{ak}	Anode to cathode, all other elements (electrodes) earthed
C_g	Grid to all other elements (electrodes) except anode

Powers

W_a	Anode dissipation
W_g	Grid dissipation
W_0	Output power

Resistances

R_a	External anode resistance
$R_{a\sim}$	External anode A.C. resistance or load resistance
R_{aa}	Load resistance of a push-pull amplifier (anode to anode)
R_{eq}	Equivalent noise resistance
R_f	Resistance of filament (heater)
R_g	External grid resistance
$R_{g'}$	External resistance between grid and cathode of next tube
R_i	Internal resistance
R_k	External cathode resistance
R_{kf}	External resistance between cathode and heater (filament)

Example 1: Tube PCL 86 (First letter P - heater current: 300 mA, second letter C – triode, L – output pentode, First number 8 – noval)
Triode and output pentode. Heater current: 300 mA. Base: noval

Example 2: Tube EL 34
Power Output Pentode. Heater voltage: 6.3 V. Base: octal

Example 3: Tube ABC 1 (First letter A - heater voltage: 4 V, second letter B – double diode, C – triode , First number 1 – miscellaneous)
Double diode and triode. Heater voltage: 4 V

Voltage stabilizers
The type designation consists of**:**

Number followed by a Capital LETTER, a figure (in some cases by a second capital letter)

Number – Burning Voltage	**First letter: Current range**
	A – max. 10 mA
	B – max. 22 mA
	C – max. 40 mA
	D – max. 100 mA
	E – max. 200 mA

Example 1: Tube 85A2 (Burning voltage 85 – 85 V, first letter A – max. 10 mA)
Vm (Ik = 5.5mA) = 83 - 87V

Example 2: Tube 150B2 (Burning voltage 150 – 150 V, first letter B – max. 22 mA)
Vm (Ik = 10mA) = 146 - 154V

GENERAL RECOMMENDATIONS

Absolute maximum ratings are limiting values of operating and environmental condition applicable to any electronic device of a specified type as defined by its published data, and should not be exceeded under the worst probable conditions.

Design maximum ratings are limiting values of operating and environmental condition applicable to a **bogey** electronic device* of a specified type as defined by its published data, and should not be exceeded under the worst probable conditions.

Design center ratings are limiting values of operating and environmental condition applicable to a bogey electronic device* of a specified type as defined by its published data, and should not be exceeded under average conditions.

***A bogey tube is a tube whose characteristics have the published nominal value for the type.**

Electrode voltage
All electrode voltages are given with respect to cathode.

- V_{a0}, V_{g20}... (limiting values of electrode voltage).
  These values are continuously permitted at zero anodes current and with cold cathode.

  They are also permitted as peak voltage during operation when a D.C. voltage in combination with a superimposed A.C. voltage is present at the electrode provided that the peak value coincides with approx. zero current.

- V_a, V_{g2}...
  These values are D.C. components of the electrode voltages and are continuously permitted.

Electrode current
- The limiting values I_a, I_{g2}, I_k, etc. are the D.C. components of the electrode currents averaged over 50 ms period.
- Limiting values could not be exceeded by more than 10 % under the worst probable conditions.

Electrode dissipation

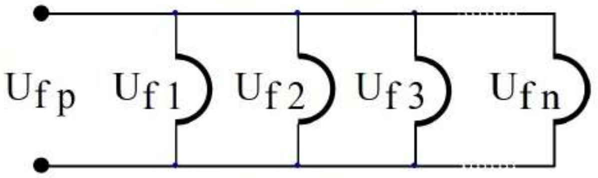
- The limiting values $\mathbf{W_a}$, $\mathbf{W_{g2}}$, etc, are the average values, obtained by averaging over any 1s period.

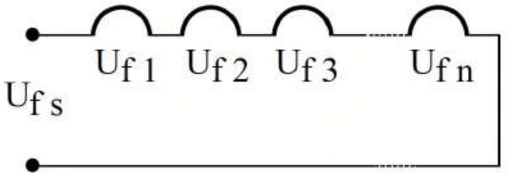
- If not otherwise indicated the quoted operating conditions for audio output tubes are permitted only with speech and music signals. With class B operation and speech and music excitation the quoted limiting value of anode dissipation is allowed to be exceeded by max. 10 % if measured with sinusoidal signal at 2 / 3 of maximum drive.
  Spread and variation in the electrode dissipation of audio output tubes should be restricted so that with bogey tubes the specified design center limiting values are not exceeded by more than 20 % under the worst probable conditions.

Heater circuit

- The maximum deviation of the heater voltage should not exceed ± 15 % (design max. value).
  This condition will be fulfilled when the mains voltage fluctuates by ± 10 % and a ordinary transformer is used.

- Heaters of several tubes can be connected in parallel or in series:





$$U_{fp} = U_f$$

$$U_{fs} = \sum_{i=1}^{n} U_{fi}$$

 # Double triode heaters can be connected in series or in parallel.

In some types of double triodes, the heaters have a common point (pin), so the heaters can be connected in series or in parallel.

- Power supply (voltage source) for the vacuum tube heater (filaments) can be AC or DC.

 # When the heater is supplied from a DC voltage source, the condition relating to the published limiting value U_{kf} must be fulfilled.
This means that the electric potential difference (DC voltage) between the cathode and the heater must belower than the limiting value U_{kf}.

Example:
In the amplifier circuit, Fig. A, the electrical potential difference (DC voltage U_{k2}) between the cathode and the ground of one (V2) of the triode systems is + 160 V - very close to the limiting value U_{kf} = 180 V of the tube used.
A DC power supply circuit used for the heaters of the tube can be designed as shown in Fig. B.

The voltage between the cathode and the heater of the triode V1 is:

$(75.9 - {}^{*}U_k = +1)$ V = ***74.9 V*** $< U_{kfmax} = 180$ V; $U_k = +1$ V.

The voltage between the cathode and the heater of the triode V2 is: (160 - 75.9) V = ***84.1 V*** $< U_{kfmax} = 180$ V

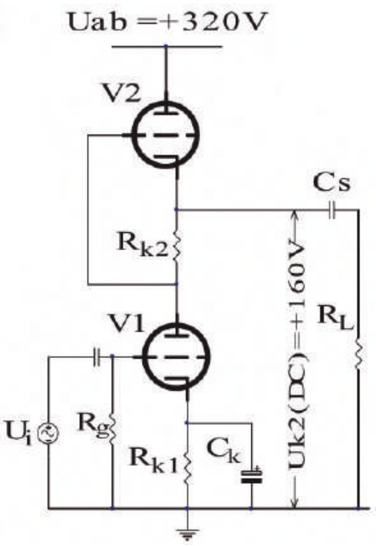


Fig. A

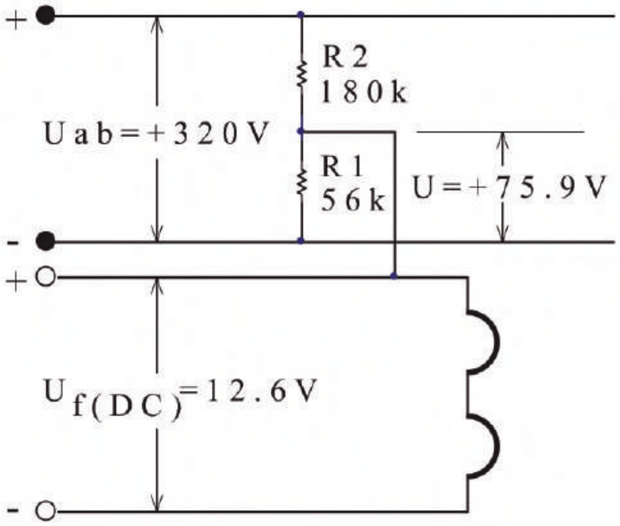


Fig. B

Microphony

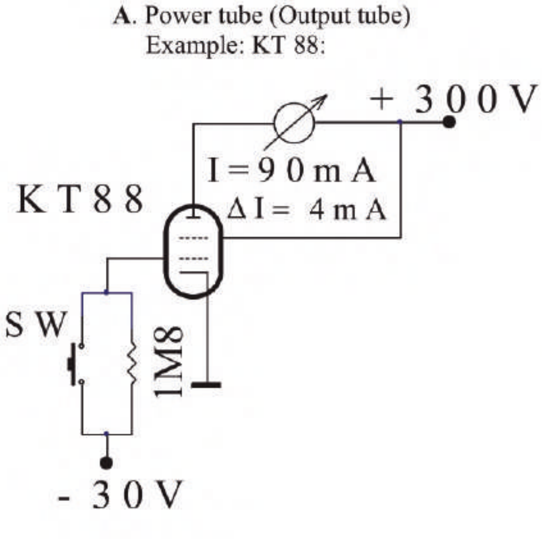
Whenever a tube is subjected to mechanical vibrations (vibrations produced by the transformers, the vibrations produced by the loudspeakers (sound), …), unwanted disturbances occur in the output signal of the tube. The effects of these disturbances depend on the mechanical characteristics of the chassis of the amplifier, the layout of electronic components, but also on the sensitivity of the tube itself to mechanical vibrations as well as on the configuration of the amplifier circuit and the conditions of application of the tube in the amplifier circuit (e.g.: amplification of the tube).

A few examples of useful tube testing tools that are not often used in practice

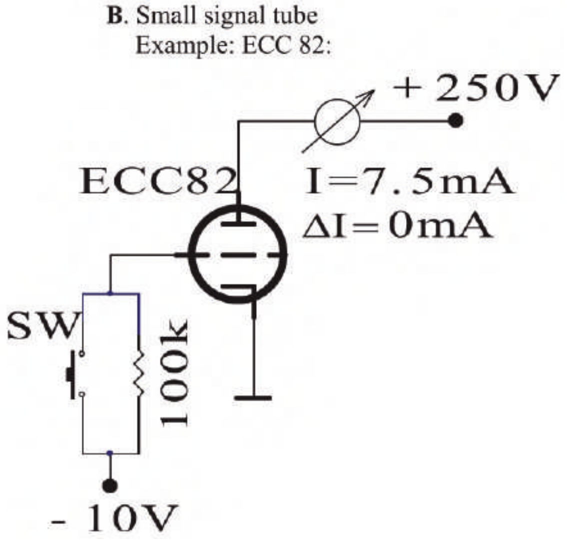
(Of course, for detailed testing of tube characteristics, the use of professional tube testers is recommended).

A simple method for testing tube vacuum:

Vacuum quality can be tested using a circuit of a very simple configuration:



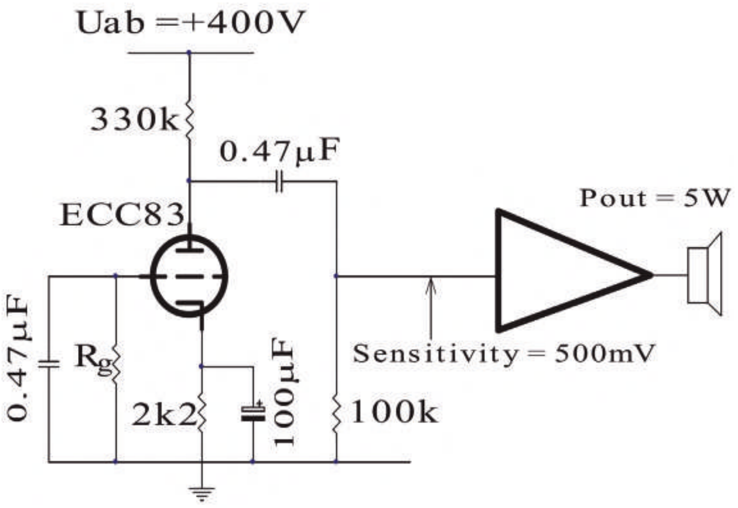
Anode current should not change more than 5% by pressing the SW push button switch.



Anode current must stay unchanged by pressing the SW push button switch.

Microphony test:
Small signal tube

Example: ECC 83:

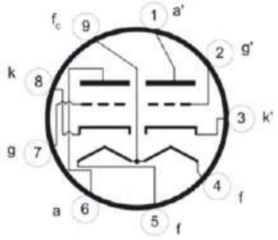


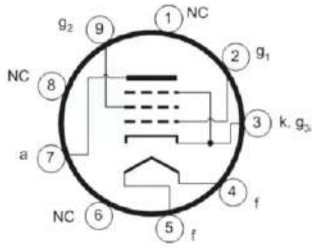
Knock (tap) the tube with a rubber hammer and listen to the microphony.

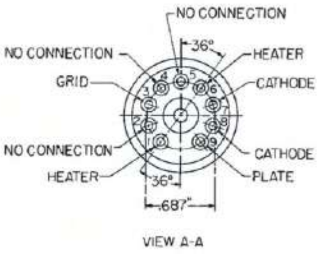
In the history of the development of vacuum tube constructions and manufacturing technologies, there have been many different tube designs related to the shapes and dimensions of the glass balloons, the shapes and dimensions of the bases and contact pins of the tubes, the layouts of the contact pins…

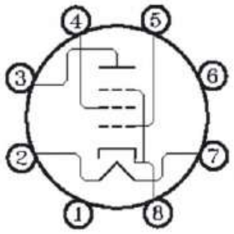
For practical reasons, the basic mechanical and electrical characteristics of tubes, their numerical, dimensional, symbolic and graphic nomenclature as well as some other rules are regulated by technical standards.

One of the rules is: the layout of the contact pins and their connection with the tube electrodes is always drawn as a **bottom view** of the tube, for example:

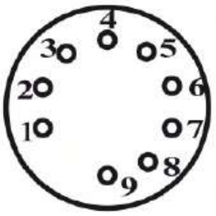








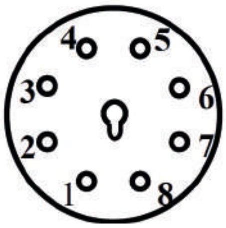
Depending on the type of tube base, the most commonly used tube types today are:
NOVAL, OCTAL, 7 pins miniature, MAGNOVAL, and specific types (WE300B, …).



NOVAL (9 pins) base, standard*:*

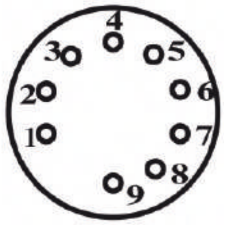
8 x 36^0 then 72^0 between pins 1 and 9, 1.016 mm diameter pins in a 11.89 mm diameter circle.

OCTAL (8 pins) base, standard*:*



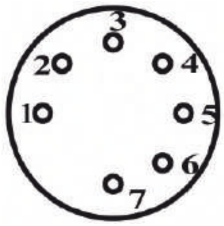
8 x 45^0, 2.36 mm diameter pins in a 17.45 mm diameter circle.

MAGNOVAL (9 pins) base, standard*:*



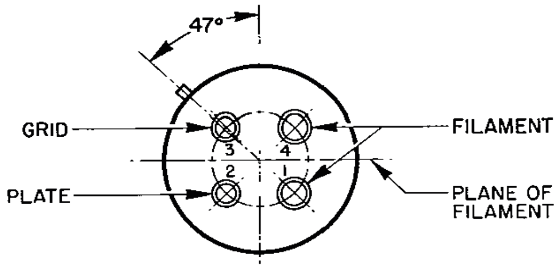
8 x 36^0 then 72^0 between pins 1 and 9, 1.27 mm diameter pins in a 17.45 mm diameter circle.

7 pins miniature base, standard*:*



6 x 45^0 then 90^0 between pins 1 and 7, 1.016 mm diameter pins in a 9.53 mm diameter circle.

U4A (4pins) base, standard*:*



11.9mm rectangle with thicker pins 1 and 4.
2 pins of 3.2 mm diameter and
2 thicker pins of
4.0 mm diameter.

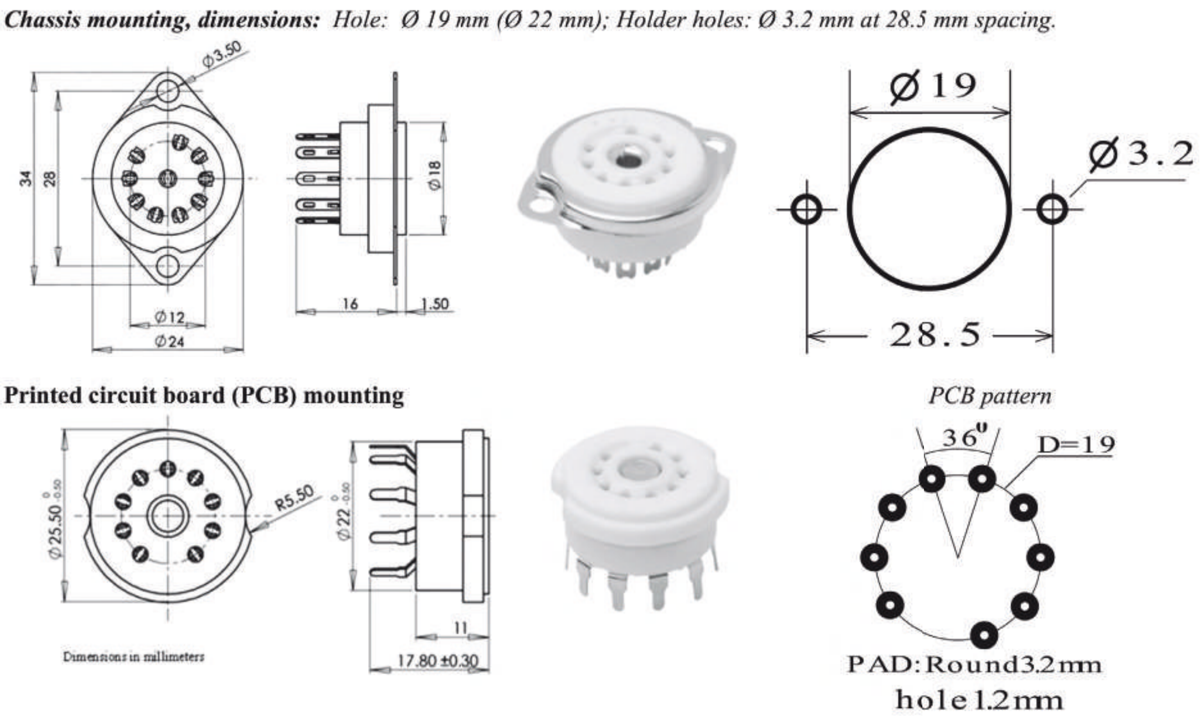
 # Note: ***BOTTOM VIEW***

In practice, each tube is placed in its corresponding type of ***socket****.*

 # Note: Most manufacturers produce tube sockets as follows:

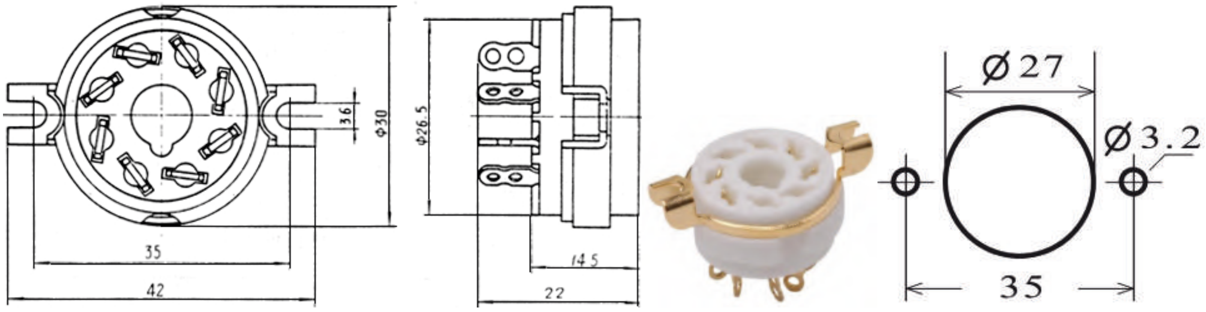
9-pin NOVAL socket

Chassis mounting, dimensions: *Hole: Ø 19 mm (Ø 22 mm); Holder holes: Ø 3.2 mm at 28.5 mm spacing.*



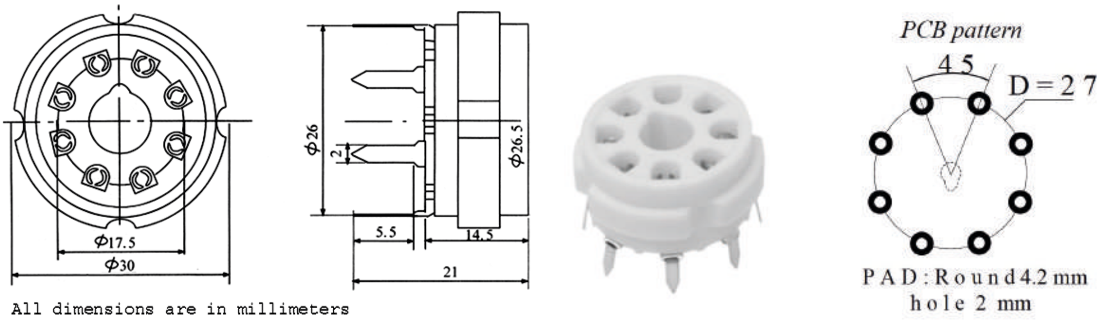
8-pin OCTAL socket

Chassis mounting, dimensions: *Hole: Ø 26.5 mm; Holder holes: Ø 3.2 mm at 42 mm spacing.*

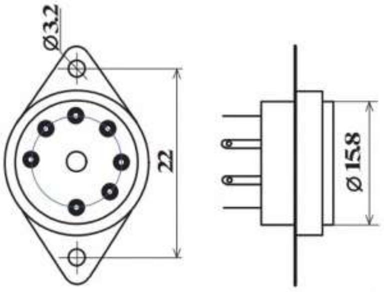


 # Note: *In general, the socket can be mounted on the upper or under side of the chassis.*
When using some types of OCTAL base tubes, it is necessary to mount the sockets on the upper side of the chassis because the diameters of the base of the tubes are larger than the distance between the holes for the holders.

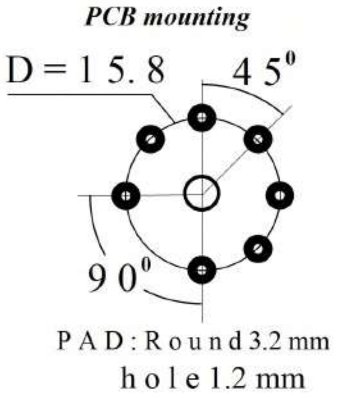
Printed circuit board (PCB) mounting



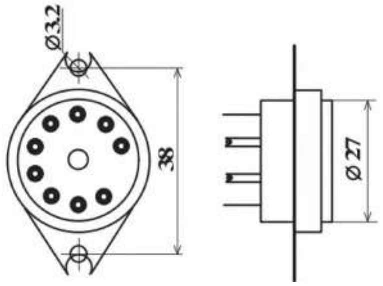
7-pin miniature socket



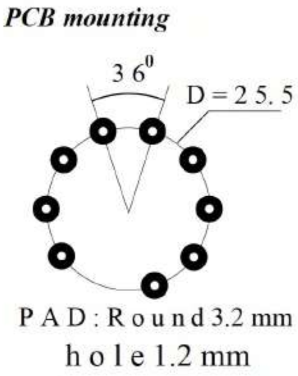
Chassis mounting
Dimension:
Hole: Ø 15.8 mm
Holder holes: Ø 3.2 mm
at 22 mm spacing.



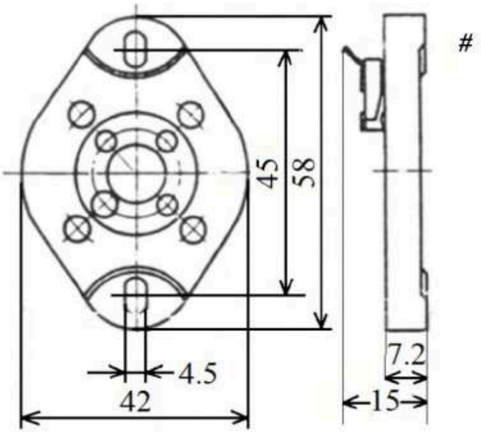
9 pin MAGNOVAL socket

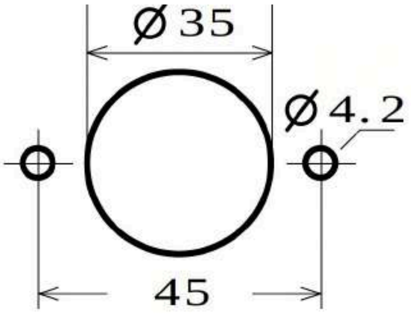


Chassis mounting
Dimension:
Hole: Ø 27 mm
Holder holes: Ø 3.2 mm
at 38 mm spacing.



***U4A socket**

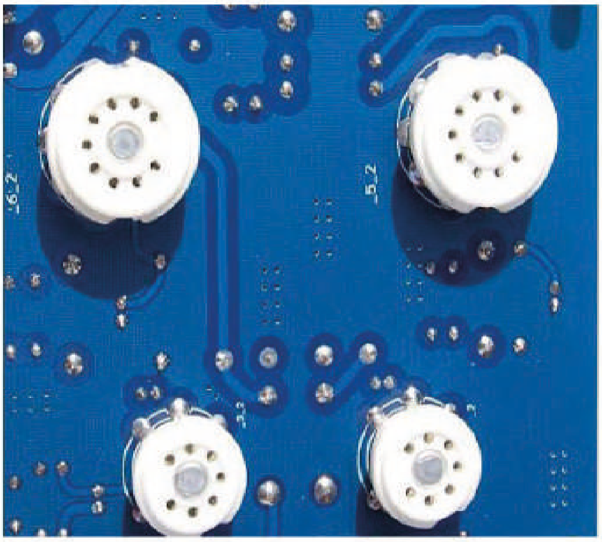




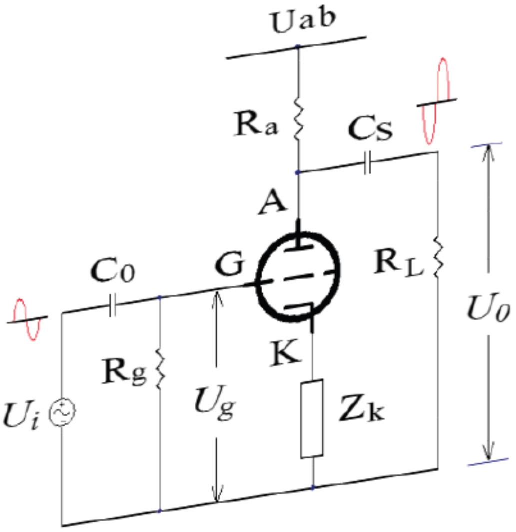
***# Note:** The 300B socket must be mounted at the **underside** of the chassis.*

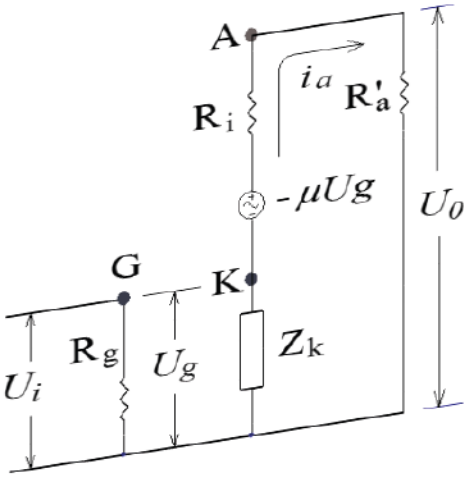
*# **Note:** The PCB-mount socket can be fitted at either side of the PCB (usually on the component side, but equally at the copper track side).*

** Keep in mind that the pattern is **mirrored** when the socket is swapped over from one side of the PCB to the other.*



Chapter 4 • Vacuum Tube Electronic Circuits





4.1 GROUNDED CATHODE (anode follower)

BASIC CIRCUIT

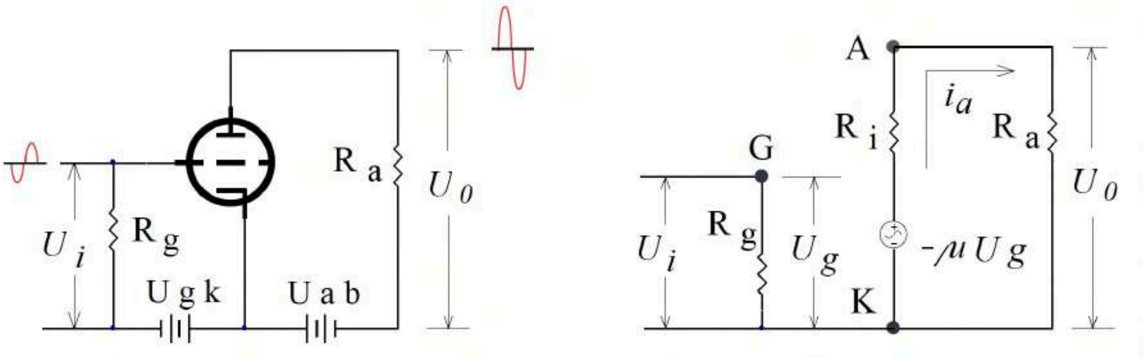


Fig. 4-01

Analyzing the circuit shown in Fig. 4-01 (and using equations: 3-07, 3-08, 3-09 from Chapter 3) the anode current is:

$$i_a = \frac{-\mu \times U_g}{R_i + R_a} .$$

Considering that $\boldsymbol{U_g = U_i}$ and that the output voltage is $U_0 = i_a \times R_a$ i.e. ($i_a = U_0 / R_a$), the above equation can be written in the following form:

$$\frac{U_0}{R_a} = \frac{-\mu \times U_i}{R_i + R_a} \quad \rightarrow \quad \frac{U_0}{U_i} = -\frac{\mu \times R_a}{R_i + R_a}$$

The amplification of the grounded cathode circuit is:

$$\boldsymbol{A} = -\mu \times \frac{R_a}{R_i + R_a} \qquad (401.1)$$

Sign " **–** " in amplification equation means that input and output voltage are **opposite in phase**.

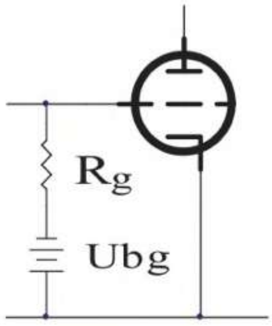
Output impedance (Z_{out}):
Analyzing the circuit from the Fig. 4-01, the output impedance is the parallel connection of R_i and R_a:

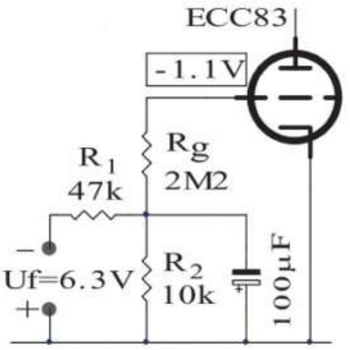
$$\boldsymbol{Z_{out}} = \frac{R_i \times R_a}{R_i + R_a} \qquad (401.2)$$

Input impedance (Z_{in})
The input impedance of the grounded cathode amplifier is the parallel connection of the grid resistor Rg and the input capacitance $\boldsymbol{C_{in}}$ (the tube grid resistance is very high and can be neglected).

$$\boldsymbol{Z_{in}} = R_g \parallel C_{in} \qquad (401.3)$$

$\boldsymbol{C_{in}}$ is the Muller capacitance: $\boldsymbol{C_{in}} = C_{gk} + C_{ga} \times (A + 1)$





The required bias voltage (**fixed bias**) of the tube can be made by inserting a voltage source into the grid circuit: *It is not necessary that U_{bg} must be a battery, it can be any stable well-filtered voltage source, (most often a well-filtered rectifier circuit is used, but it can be used for example and the regulated DC tube heater power supply).*

***Example**:*
ECC83 grounded cathode circuit design requirement: $U_g = -1.1V$.

The DC heater power supply Uf = 6.3V is available:

$$U_g = 1.1V = \frac{R_2}{R_1+R_2} \times U_f \quad ; \quad R_1 = 47k\Omega \; and \; R_2 = 10k\Omega$$

GROUNDED-CATHODE AMPLIFIER WITH IMPEDANCE IN THE CATHODE CIRCUIT

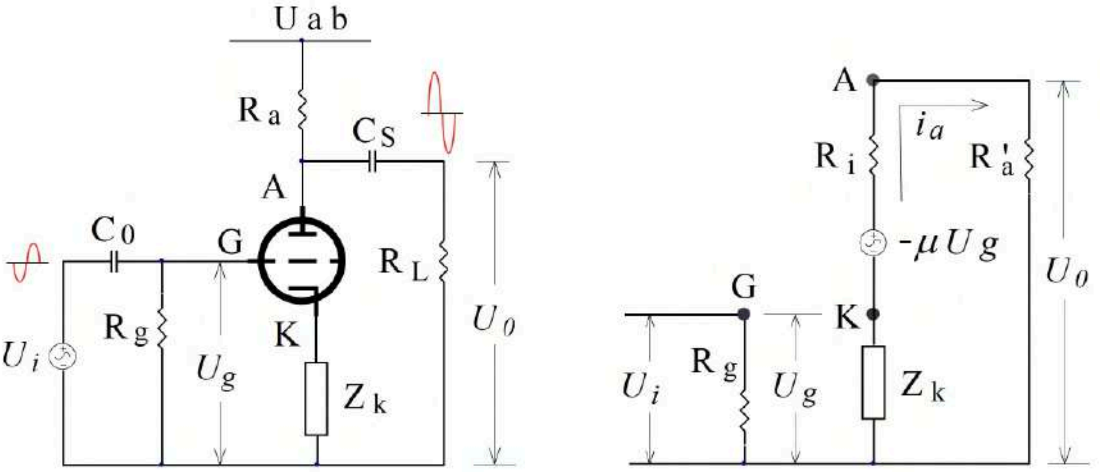


Fig. 4-02

Fig. 4-02 (equations: 3-07, 3-08, 3-09; Chapter 3): anode current is $i_a = \frac{\mu \times U_g}{R_i+R'_a+Z_k}$ and $R'_a = R_a \| R_L = \frac{R_a \times R_L}{R_a+R_L}$.

Grid – cathode voltage is equal to the input voltage plus the AC voltage across $\boldsymbol{Z_k}$:

$$\boldsymbol{U_g} = U_i + i_a \times Z_k \tag{401.4}$$

$$\boldsymbol{i_a} = \frac{-\mu \times U_g}{R_i + R'_a + Z_k} = \frac{-\mu \times (U_i + i_a \times Z_k)}{R_i + R'_a + Z_k} = \frac{-\mu \times U_i}{R_i + R'_a + (\mu + 1) \times Z_k}$$

Also, $\boldsymbol{U_0} = i_a \times R'_a \quad \rightarrow \quad \boldsymbol{i_a} = \frac{U_0}{R'_a}$, , hence:

$$\frac{U_0}{R'_a} = \frac{-\mu \times U_i}{R_i + R'_a + (\mu + 1) \times Z_k} \quad \rightarrow \quad \frac{U_0}{U_i} = \frac{-\mu \times R'_a}{R_i + R'_a + (\mu + 1) \times Z_k}$$

The amplification of the grounded cathode amplifier with impedance Z_k in the cathode circuit is:

$$\boldsymbol{A} = -\frac{\mu \times R_a^{'}}{R_i + R_a^{'} + (\mu+1) \times Z_k} \tag{401.5}$$

The impedance Z_k in the cathode circuit acts as a Negative Feed Back - [the internal tube resistance R_i increases to the value $R_i+(\mu+1)xZ_k$], and reduces the amplification of the grounded cathode amplifier.

The amplification equation of the circuit with NFB applied to Equation (401.5) is:

$$\boldsymbol{A} = \frac{A_0}{1+\beta_r \times A_0} \;; \qquad \boldsymbol{\beta_r} = \frac{Z_k}{R_a'} + \frac{Z_k}{\mu \times R_a'}, \; and \; \boldsymbol{A_0} = -\mu \times \frac{R_a'}{R_a'+R_i}$$

Output impedance (Z_{out})

Considering that the internal tube resistance R_i is increased to the value $R_i + (\mu + 1) \times Z_k$ (the amplification equation (401.4)) the output impedance of the grounded cathode amplifier with impedance Z_k in the cathode circuit is the parallel connection $R_i + (\mu + 1) \times Z_k$ and resistor R_a of the anode circuit:

$$\boldsymbol{Z_{out}} = R_i' \| R_a = \frac{R_i' \times R_a}{R_i' + R_a} \;; \qquad \boldsymbol{R'_i} \text{ is: } \boldsymbol{R_i'} = R_i + (\mu+1) \times Z_k \tag{401.6}$$

In order to determine the optimum loading of the amplifier stage or the input impedance of the next amplifier stage acting as the load of the grounded cathode amplifier stage, it is very important to determine and calculate the output impedance of the grounded cathode amplifier stage.

Input impedance (Z_{in})

The input impedance of the grounded cathode amplifier is the parallel connection of the grid circuit resistor R_g and the input capacitance C_{in} of the tube (the grid resistance is very high and can be neglected).

$$\boldsymbol{Z_{in}} = R_g \parallel C_{in} \tag{401.7}$$

C_{in} is the Miller capacitance: $\boldsymbol{C_{in}} = C_{gk} + C_{ga} \times (A+1)$

High - frequency response:

The effect of the input circuit to the frequency response of the grounded cathode amplifier at high frequencies is determined by the output resistance (R_0) of the signal source that drives the grounded cathode amplifier circuit and the input capacitance of the grounded cathode amplifier (C_{in}).

The cutoff frequency (a frequency for which the voltage amplification is -3dB of the nominal value) can be calculated as follows:

$$\boldsymbol{f_{H(-3dB)}} = \frac{1}{2 \times \pi \times R_0 \times C_{in}} \tag{401.8}$$

(reminder: C_{in} is the Miller capacitance):

$$\boldsymbol{C_{in}} = C_{gk} + C_{ga} \times (A+1)$$

Low - frequency response:

The effect of the input circuit to the response at low frequencies is determined by C_0 and Rg.
The cutoff frequency can be calculated as follows:

$$\boldsymbol{f_{L(-3dB)}} = \frac{1}{2 \times \pi \times R_g \times C_0} \tag{401.9}$$

FIRST OPTION: $\boldsymbol{Z_k = 0}$

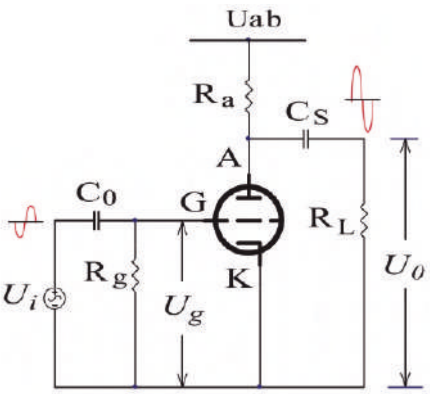


Fig. 4-03

The cathode is connected to the ground.

The grid voltage (the voltage across $\boldsymbol{R_g}$ caused by the flow of a very low current ($\boldsymbol{I_g}$) of the grid) is very low. In practice, high resistance $\boldsymbol{R_g}$ is used.

This type of circuit is used as a low input signal amplifier.

Amplification:
Since the resistor $\boldsymbol{R_a}$ of the anode circuit is actually connected in parallel with the load resistor $\boldsymbol{R_L}$ (impedance of $\boldsymbol{C_s}$ is very low and can be neglected) the effective anode resistance $\boldsymbol{R'_a}$ is:

$$\boldsymbol{R'_a} = R_a \| R_L = \frac{R_a \times R_L}{R_a + R_L}$$

*# **Note: Effective anode resistance R'**$_a$ is of interest in **AC** circuit analysis (**R**$_a$ is used in **DC** analysis).*

Amplification (*A*) $A = -\frac{U_{out}}{U_{in}} = -\mu \times \frac{R'_a}{R_i + R'_a}$ (401.10)

Input impedance (Z_{in}): $\boldsymbol{Z_{in}} = R_g \| C_{in}$ where $\boldsymbol{C_{in}} = C_{gk} + C_{ga} \times (A+1)$

Output impedance (R_{out}): $\boldsymbol{R_{out}} = R_i \| R_a = \frac{R_i \times R_a}{R_i + R_a}$ (401.11)

SECOND OPTION: $\boldsymbol{Z_k = R_k}$

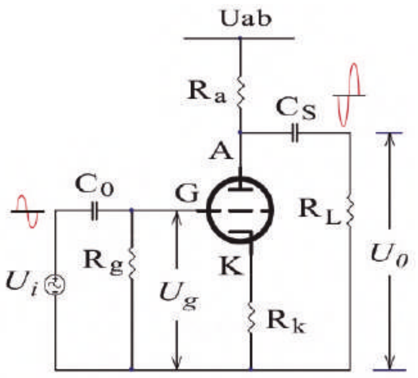


Fig. 4-04

The required bias voltage (**automatic bias voltage**) of the tube can be made by inserting a resistor R_k in a cathode circuit:

$$-\mathbf{U_g} = \mathrm{R_k} \times \mathrm{I_a}$$

The amplification (A) of the circuit:
The automatic bias (resistor $\boldsymbol{R_k}$ in a cathode circuit) acts as an NFB [the internal resistance of the tube $\boldsymbol{R_i}$ is increased to the value $\boldsymbol{R_i+(\mu+1)} \times \boldsymbol{R_k}$], and reduces the amplification of the grounded cathode amplifier:

$$A = -\mu \times \frac{R'_a}{R'_a + R_i + (\mu+1) \times R_k} \qquad (401.12)$$

The amplification with NFB (Equation (401.12)) is: $A = \frac{A_0}{1 + \beta_r \times A_0}$ *where:* $\beta_r = \frac{R_k}{R'_a} + \frac{R_k}{\mu \times R'_a}$

 # In practice $\frac{R_k}{R'_a} \gg \frac{R_k}{\mu \times R'_a}$ *, therefore, the* $\boldsymbol{\beta_r}$ *is:* $\beta_r \approx \frac{R_k}{R'_a}$

Finally, the amplification A is: $A=\dfrac{A_0}{1+\dfrac{R_k}{R_a^{'}}\times A_0}$, ***A₀*** *is:* $A_0=-\mu\times\dfrac{R_a^{'}}{R_a^{'}+R_i}$ (401.13)

Impedance (***Rout*** **):** $R_{out}=R_i^{'}\|R_a=\dfrac{R_i^{'}\times R_a}{R_i^{'}+R_a}$ (401.14)

R'ᵢ is: $R_i^{'}=R_i+(\mu+1)\times R_k$ (401.15)

THIRD OPTION: $Z_k=R_k\|C_k$ **Bypassed cathode resistor**

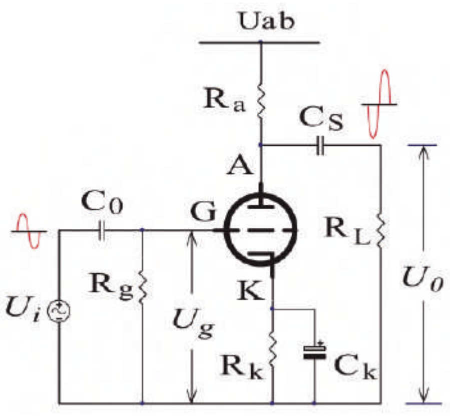


Fig. 4-05

Via the capacitor ***Cₖ ,*** the AC signal by bypasses the cathode resistor ***Rₖ*** to ground, so that no AC current (signal current) flows through the cathode resistor (there is no AC voltage across the R_k in equation (401.4)), while the DC voltage across the R_k (grid bias voltage) remains unchanged. The effect of NFB (caused by the impedance in the cathode circuit which reduces the amplification and increases the output impedance) can be partially or fully eliminated.

The impedance in the cathode circuit is the parallel connection of ***Rₖ*** and ***Cₖ***:

$Z_k=\dfrac{R_k}{1+j2\pi f R_k C_k}$, and amplification is:

$$A=-\mu\times\frac{R_a}{R_a++R_i+(\mu+1)\times Z_k} \quad \text{or:} \quad A=-\mu\frac{R_a}{R_i+R_a}\sqrt{\frac{1+(2\pi f R_k C_k)^2}{[1+\frac{R_k(\mu+1)}{R_i+R_a}]^2+(2\pi f R_k C_k)^2}}$$

If the capacitor ***Cₖ*** is large enough that (μ+1) x Z_k is small compared to R_i + R_a (i.e. a fully bypassed cathode resistor), the **amplification (A)** is:

$$A=A_0=-\mu\times\frac{R_a^{'}}{R_a^{'}+R_i} \qquad (401.16)$$

$$R_a^{'}=R_a\|R_L=\frac{R_a\times R_L}{R_a+R_L}$$

Output impedance (R_{out}):

$$R_{out}=R_i\|R_a=\frac{R_i\times R_a}{R_i+R_a} \qquad (401.17)$$

Where ***Rᵢ*** is the internal resistance of the tube.

Cathode impedance (*Rkeq*):

 # Cathode impedance is of interest in ***AC analysis*** *[****Rₖ*** *is used in DC analysis, (to calculate* ***U_g*** *, for example)].*

In graphic analysis $I_a=\dfrac{U_{ab}-U_{a-k}}{R_a}$ *. This is a straight line (**load line**) with a slope of* $\dfrac{-1}{R_a}$.

The intersection of the load line with each curve of the anode characteristic determines the anode current for the parameter of each characteristic, i.e. grid voltage U_g , (R_k is: $R_k=\dfrac{U_g}{I_a}$).

The resistance of the grounded cathode circuit (excluding the cathode resistor) seen from the cathode is:

$$R_k' = \frac{R_i + R_a}{\mu + 1} \qquad (401.18)$$

The equivalent cathode resistance is the parallel connection of the cathode resistor $\boldsymbol{R_k}$ and the resistance $\boldsymbol{R'_k}$:

$$R_{k\,eq} = R_k' \parallel R_k = \frac{\frac{R_i + R_a}{\mu + 1} \times R_k}{\frac{R_i + R_a}{\mu + 1} + R_k} \qquad (401.19)$$

 # Note
The capacitance of the capacitor $\boldsymbol{C_k}$ *affects the amplitude characteristic (frequency response) of the circuit. Thus, if the capacitance* $\boldsymbol{C_k}$ *is low, the lower frequency signal will be amplified less than the high frequency signal. (The minimum amplification of the grounded cathode amplifier circuit (at the lowest frequency) is equal to the amplification of the grounded cathode amplifier circuit with non-bypassed cathode resistor).*

The maximum amplification that can be achieved is equal to the amplification of the grounded cathode amplifier circuit with a fully bypassed cathode resistor).

The cutoff frequency can be calculated as follows:

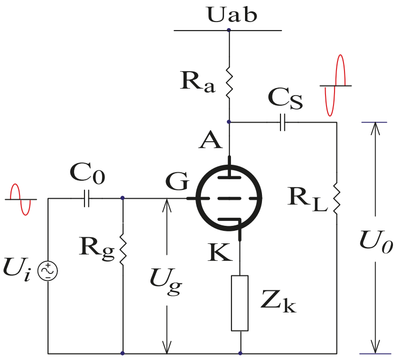
$$f_{L(-3dB)} = \frac{1}{2 \times \pi \times R_{keq} \times C_k} \qquad (401.20)$$

 # Note
The grounded cathode amplifier is the most commonly used tube circuit in practical applications. It can be said that this is a “natural” application of tubes with roots at the very beginning of the tube technique. So, when audiophiles talk about “tube sound”, they mean the sounds produced by an audio amplifier based on the grounded cathode amplifier circuit.

Analysis and design of a grounded cathode amplifier is not complicated and this tube amplifier configuration gives grate freedom to designers to achieve the desired or required characteristics of the amplifier circuit in terms of amplification, frequency response, output impedance,…. Also, very low distortion of the amplified signal can be achieved.

Grounded cathode amplifiers can be used as low signals amplifier circuits (such as input amplifier circuit), as medium signals amplifier circuits (such as driver circuits) and as power output amplifiers.

SUMMARY

<table>
<tr><th colspan="3">GROUNDED CATHODE</th></tr>
<tr><td colspan="3">
$Z_k = Z_k$
$A = -\mu \times \frac{R'_a}{R'_a+R_i+(\mu+1)\times Z_k}$
$R'_a = \frac{R_a \times R_L}{R_a+R_L}$
$R_{out} = R'_i \| R_a = \frac{R'_i \times R_a}{R'_i+R_a}$
$R'_i = R_i + (\mu+1) \times Z_k$
$Z_{in} = R_g \| C_{in}$
$C_{in} = C_{gk} + C_{ga} \times (A+1)$</td></tr>
<tr><th>$Z_k = 0$</th><th>$Z_k = R_k$</th><th>$Z_k = R_k \,\mathrm{II}\, C_k$</th></tr>
<tr><td>$A = -\mu \times \frac{R'_a}{R'_a+R_i}$</td><td>$A = -\mu \times \frac{R'_a}{R'_a+R_i+(\mu+1)\times R_k}$</td><td>$A = -\mu \times \frac{R'_a}{R_i+R'_a}$</td></tr>
<tr><td>$R_{out} = R_i \| R_a = \frac{R_i \times R_a}{R_i + R_a}$</td><td>$R_{out} = R'_i \| R_a = \frac{R'_i \times R_a}{R'_i + R_a}$
$R'_i = R_i + (\mu+1) \times R_k$</td><td>$R_{out} = R_i \| R_a = \frac{R_i \times R_a}{R_i + R_a}$
$R_{keq} = R'_k \| R_k = \frac{\frac{R_i + R_a}{\mu+1} \times R_k}{\frac{R_i + R_a}{\mu+1} + R_k}$
$R'_k = \frac{R_i + R_a}{\mu+1}$
$C_k = \frac{1}{2 \times \pi \times R_{keq} \times f_{L(-3dB)}}$</td></tr>
</table>

Main characteristics:

- high input impedance
- medium - to - low output impedance
- relatively high amplification
- good frequency response
- input and output signals are opposite in phase

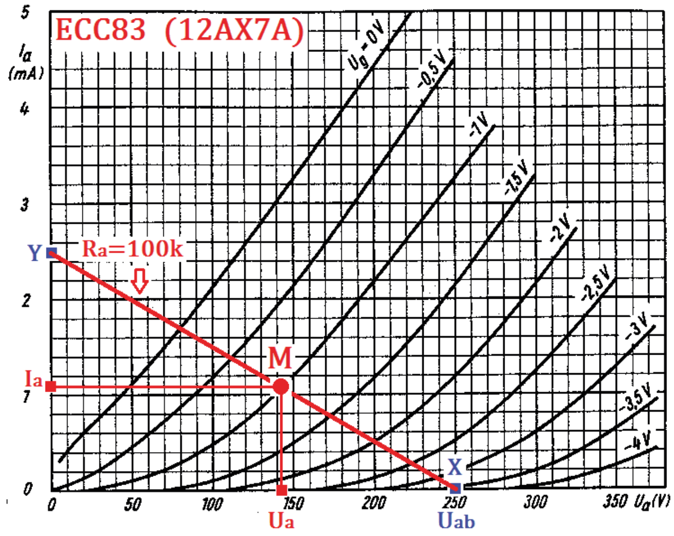


Fig. 4-06

Example 401.1:

Low-signal amplifier. Grounded cathode (anode follower) amplifier.

Calculate: ***A_0 , A, R_k***

(1 / 2 ECC83, i.e. one triode).

Determine tube working point: **M (U_a, I_a)**.

Load: grid resistor of the next stage

R_L = R_g = 1MΩ.

U_{ab} = 250V

R_a = 100 kΩ

U_g = -1V

- **ECC83**

  Reference data (one triode):

  I_a = 1.25 mA

  μ = 100

  R_i = 62.5kΩ

1. Calculation of R'_a : $R_a' = R_a \| R_L = \frac{R_a \times R_L}{R_a + R_L} = \frac{100k \times 1000k}{100k + 1000k} = 90.91k$

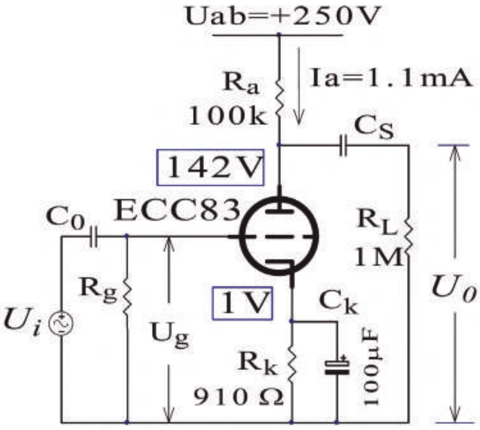
2. Calculation of A_0 : $A_0 = -\mu \times \frac{R_a'}{R_a' + R_i} = -100 \times \frac{90.91k}{90.91k + 62.5k} = -59.26$

3. Drawing the Load line:

- *Draw the point* ***X*** *on the* ***U_a axis****:* ***U_a = U_{ab} = 250 V***
- *Draw the point* ***Y*** *on the* ***I_a axis****:* ***U_{ab} / R_a = 250 V / 100 kΩ = 2.5 mA***
- *Draw a straight line passing through points* ***X*** *and* ***Y*** *– this is the* ***load line.***
- ***Load line*** *intersects the* ***U_g = - 1 V characteristic*** *at point* ***M****. Coordinate of point M:* ***M (142 V, 1.1 mA),*** *i.e.* ***U_a = 142 V, I_a = 1.1 mA***

4. Calculation of R_k : $R_k = \frac{U_g}{I_a} = \frac{1}{0.0011} = 909\,\Omega$ *(Standard:* ***R_k = 910 Ω****)*

5. Calculation of A: $A = \frac{A_0}{1 + \frac{R_k}{R_a'}} = \frac{59.26}{1 + \frac{0.91}{90.91} \times 59.26} = 37.19$



Note

In practical application (mostly as the second stage of the preamplifier) this circuit is used very often, with a few changes regarding power supply voltage

(U_{ab} = 260V) and R_k (R_k = 1kΩ).

Example 401.2:

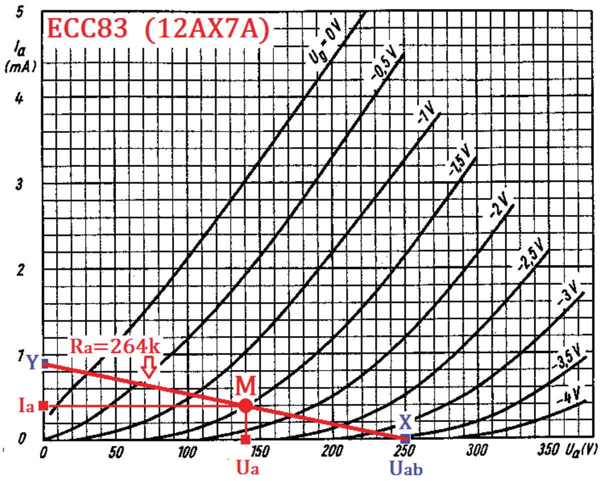


Fig. 4-07

- Design an anode follower amplifier
  A_0 = 77
  Input signal withstand: **1V_{RMS}**
- Available power supply: U_{ab} = 250V.
- Tube
  1 / 2 12AX7(ECC83) [one triode].
- The amplifier stage is loaded with a grid resistor of the next stage: R_L= R_g= 1MΩ.

- **ECC83**
  Reference data (one triode):
  I_a = 1.25 mA
  μ = 100
  R_i = 62.5kΩ

1. ***Calculation of R'$_a$:*** $A_0 = -\mu \times \frac{R'_a}{R'_a + R_i} \rightarrow R'_a = \frac{R_i \times A_0}{\mu - A_0} = \frac{62.5\,k \times 77}{100-77} = 209\,k\Omega$

2. ***Calculation of R$_a$:*** $R'_a = R_a \| R_L = \frac{R_a \times R_L}{R_a + R_L} \rightarrow R_a = \frac{R'_a \times R_L}{R_L - R'_a} = \frac{209\,k \times 1000\,k}{1000\,k - 209\,k} = 264\,k\Omega$

3. ***Drawing of the Load line:***

- *Draw the point **X** on the U_a axis: **U_a = U_{ab} = 250 V** (available voltage source).*
- *Draw the point **Y** on the I_a axis:* $\frac{U_{ab}}{R_a} = \frac{250\,V}{264\,k} = 0.95\,mA$
- *Draw a straight line passing through points X and Y – this is the **load line**.*
- *The condition to withstand the input signal of **1V_{RMS}** is fulfilled if:*

  $\boldsymbol{U_{g-k}} \geq U_{Peak\ in} = U_{RMS\ in} \times \sqrt{2} = 1V \times 1.41 = \mathbf{1.41\ V}$

  *(Choice: **U_{g-k} = - 1.5 V**)*
  *The quiescent point is located at the intersection of the load line and the anode characteristic **U_g = - 1.5 V** i.e. coordinate of point M is: **M (140 V, 0.4 mA)**.*
  *Tube operating conditions (quiescent point): **U_a = 140 V, U_{g-k} = - 1.5 V** and **I_a = 0.4 mA.***

4. ***Calculation of cathode resistor R$_k$:*** $R_k = \frac{U_k}{I_a} = \frac{1.5\,V}{0.4\,mA} = 3.75\,k\Omega$ *(Standard: **3.9 kΩ**)*

5. ***Cathode impedance (R$_{keq}$):*** $R'_k = \frac{R_i + R_a}{\mu + 1} = \frac{62.5\,k + 264\,k}{100+1} = 3.23\,k\Omega$

$$R_{keq} = R'_k \| R_k = \frac{\frac{R_i + R_a}{\mu+1} \times R_k}{\frac{R_i + R_a}{\mu+1} \times R_k} = 1.76\,k\Omega$$

6. ***Calculation of C$_k$:***

 # Using equation (401.20): $f_{L(-3dB)} = \frac{1}{2 \times \pi \times R_{keq} \times C_k} \rightarrow C_k = \frac{1}{2 \times \pi \times R_{keq} \times f_{L(-3dB)}}$

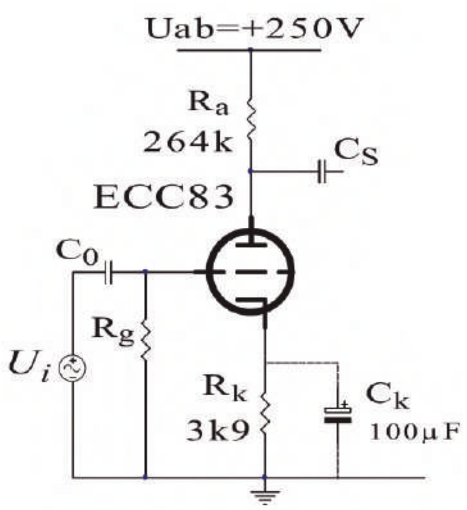
If the required $f_{L(-3dB)}$ *is, for example, 1Hz:*

$$C_k = \frac{1}{2 \times \pi \times R_{keq} \times f_{L(-3dB)}} = \frac{1}{2 \times 3.14 \times 1760 \times 1} = 90\,\mu F$$

(Standard: ***C_k = 100 µF****)*

7. Recalculation of A_0: $A_0 = -\mu \times \frac{R_a'}{R_a' + R_i} = -100 \times \frac{209\,k}{209\,k + 62.5\,k} = 77$

8*. Without C_k , the amplification is: $A = \frac{A_0}{1 + \frac{R_k}{R_a'} \times A_0} = \frac{77}{1 + \frac{3.9\,k}{209\,k} \times 77} = 31.6$



 # Note
In practice, this circuit is very often used (mostly as an input stage of a preamplifier).

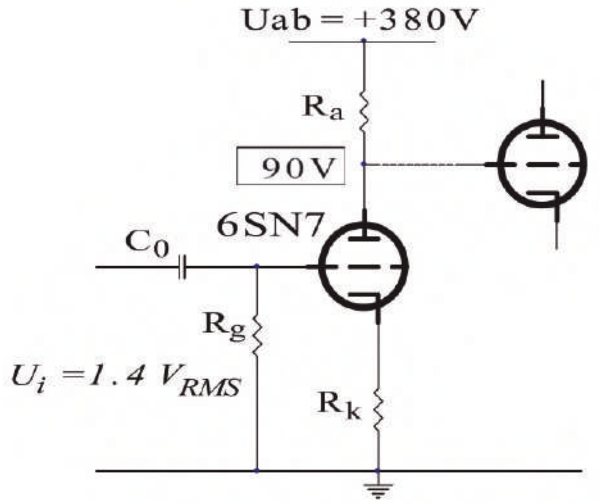


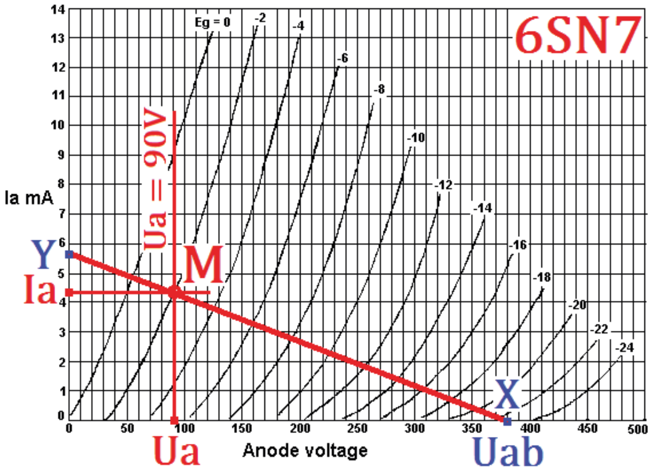
Fig.4-08

Example 401.3:
Design a grounded cathode amplifier input stage using one triode of 6SN7 (as shown in Fig. 4-08) capable of withstanding an input signal of 1.4V_{RMS} with the additional condition that U_a must be 90V (suitable for DC coupling to the next stage).

Determine: I_a, U_{gk}, R_a, R_k, A_0, A.

Tube: 1 / 2 6SN7
U_{ab} = 380V

6SN7, reference data (one triode):
R_{i1} = 7.7kΩ
$I_a = f(U_a)$ characteristics: Fig. 4-09
$\mu_1 = 20$



1. *Draw a vertical line from the point* $\boldsymbol{U_a = 90\ V}$.

2. *The condition to withstand the input signal of* $\boldsymbol{1.4\ V_{RMS}}$ *is fulfilled if:*
   $\boldsymbol{U_{gk} \geq 1.4\ V \times 1.41 = 2\ V}$
   Quiescent point *is located at the intersection of the vertical line* $\boldsymbol{U_a = 90\ V}$ *and the characteristic* $\boldsymbol{U_g = -2\ V}$ *i.e.* ***M (90 V, 4.3 mA)***.

   $\boldsymbol{U_a = 90\ V,\ I_a = 4.3\ mA}$

Fig.4-09

3. *Draw a straight line passing through the points* ***M*** *and* $\boldsymbol{U_{ab}}$ *up to the intersection with the* $\boldsymbol{I_a}$ *axis (this is the* ***load line****).*

4. *The load line intersects the* $\boldsymbol{I_a}$ *axis at point* ***Y*** $\boldsymbol{(I_{a(Y)} = 5.7\ mA)}$.

5. *Calculate* $\boldsymbol{R_a}$ *as a slope of the load line, i.e.*
   $\boldsymbol{R_a} = U_{ab} / I_{a(Y)} = 380 / 0.0057 = \boldsymbol{66667}\ \Omega$.

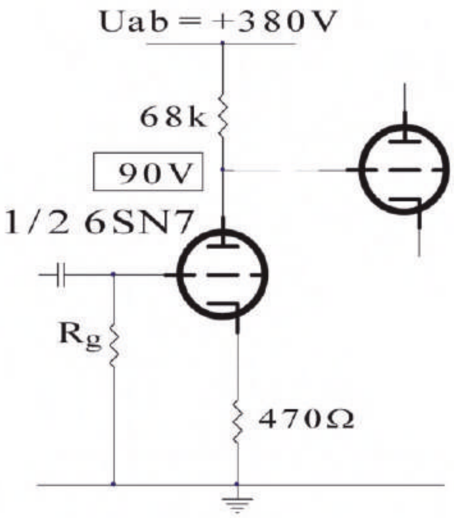
6. *Standard* R_a *:* $\boldsymbol{R_a = 68k\Omega}$.
7. *Calculation of* $\boldsymbol{R_k}$*:*
   $\boldsymbol{R_k} = U_{gk} / I_a = 2 / 0.0043 = \boldsymbol{465}\Omega$. *(Choose* $\boldsymbol{R_k = 470\Omega}$*).*

8. *Calculation of* $\boldsymbol{A_0}$:

$$A_0 = -\mu \times \frac{R_a}{R_a + R_i} = -20 \times \frac{68\,k}{68\,k + 7.7\,k} = -17.96$$

9. *Calculation of* ***A****:*

$$A = \frac{A_0}{1 + \frac{R_k}{R_a} \times A_0} = \frac{17.96}{1 + \frac{0.47}{68} \times 17.96} = 15.97$$



 # Note
This circuit is often used in practice, for example, as an input stage of a Single Ended power amplifier with the WE300B as the output tube and with the 6SN7 second triode as a driver.

Example 401.3 / 1
Design the driver stage of the SE power amplifier with one triode of the 6SN7 (output tube: WE300B). It is necessary to drive 300B with the signal: U_{peak} = 71V (50.35V_{RMS}). The driver stage is DC coupled to the input stage discussed in Example 401.3. This means that the grid should be at an electric potential of 90V. Also, the driver stage must withstand an input signal of 5V_{RMS} and its amplification should be similar to the amplification of the input stage (around 15, say 15.2).

Calculation and design:

1. *The input signal is 5 V_{RMS} , so U_{g-k} must be* $U_{g-k}=5V_{RMS}\times\sqrt{2}=-7V$*, which means that the cathode has to be at the electrical potential:* $\boldsymbol{U_k}$ *= Ug + U_{g-k} = 90 V + 7 V =* ***97 V.***

2. *Amplification should be* ***15.2***

   Using equation $A=-\mu\times\frac{R_a}{R_i+R_a}\rightarrow R_a=\frac{A\times R_i}{\mu-A}=\frac{15.2\times 7.7k}{20-15.2}=24.38\,k\Omega$

3. *Data used in graphic analysis:*

   As the cathode is at a potential of 97 V, the new $\boldsymbol{U_{ab}}$ *(point on the U_a axis) is: 380 V – 97 V* ***= 283 V***

   The load line can be drawn using the parameters: ***R_a = 24.38 kΩ*** *and* ***U_{ab} = 283 V***
   (Load line: draw a straight line passing through the points U_{ab} (283V) of the Ua axis and the point I_a = U_{ab} / R_a =283 V / 24.38 kΩ =11.6 mA of the I_a axis).

4. ***Quiescent point*** *is located at the intersections of the* ***load line*** *and the anode* ***characteristic U_g = - 7 V***.

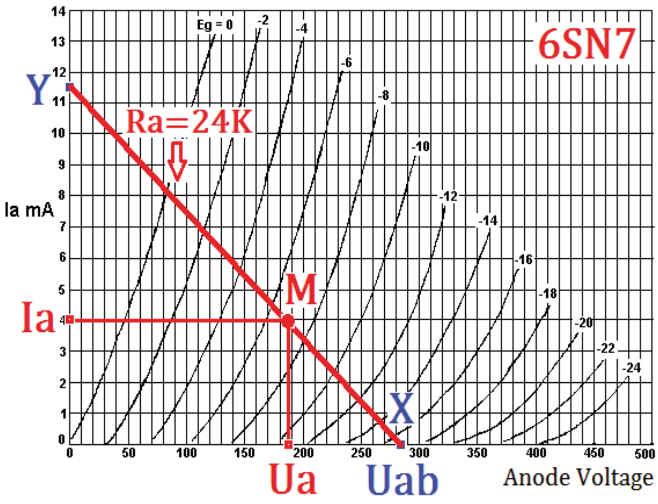
   Quiescent point: ***M (186 V, 4 mA).***

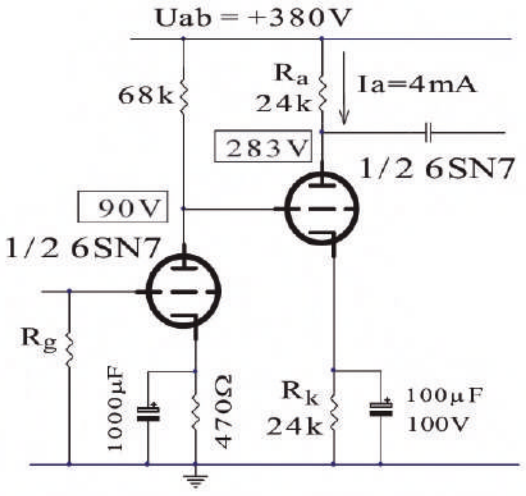
   I_a = 4 mA.

5. *Calculation of* ***R_k*** *:*

   $$R_k=\frac{U_k}{I_a}=\frac{97\,V}{4\,mA}=24.25\,k\Omega$$

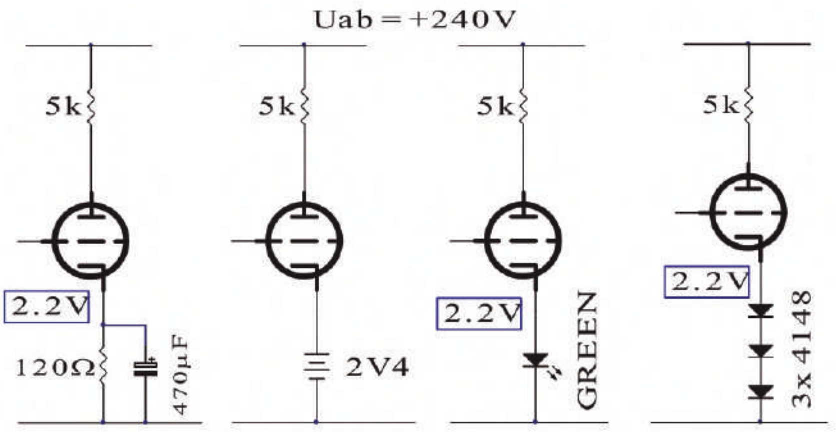
6. *Standard*: ***R_k = 24 kΩ*** *and* ***R_a = 24 kΩ***





More details on the grounded cathode amplifier:

I. The automatic bias circuit of the grounded cathode amplifier stage (anode follower) can be made using some other electronic components, and not only using resistors in the cathode circuit.



BATTERIES:
Lead acid: 2.0 V per cell
Nickel cadmium: 1.2 V per cell
Lithium-ion: 3.6 V per cell
Alkaline: 1.5 V per cell
Primary Lithium: 3 V per cell

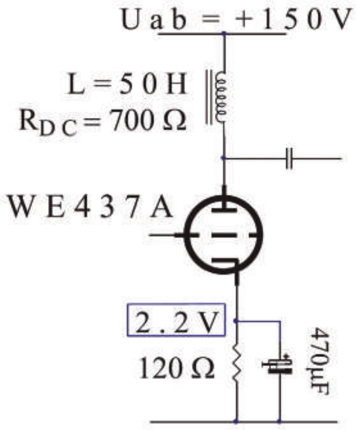
LED:

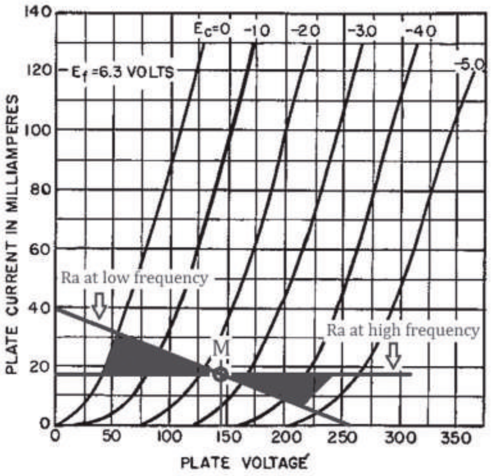
RED	***U_F = 1.7 V***
AMBER	***U_F = 2.0 V***
YELLOW	***U_F = 2.1 V***
GREEN	***U_F = 2.2 V***
BLUE	***U_F = 4.5 V***

II. Other electronic components or electronic circuits can be used as the anode load, not just resistors.

1. Inductance

The impedance of the inductance used as an anode load is not constant, it varies depending on the frequency of the signal. By using a well-defined inductance, a very high impedance of the anode load can be achieved at medium and high frequencies. Problems can occur at low frequencies when the impedance of the inductance used as the anode load can be low or insufficient.





Example:
Grounded cathode amplifier, anode load: inductance, L
Tube WE437A: μ = 41; R_i = 960 Ω
L = 50 H; R_{DC} = 700 Ω
U_{in} = 1V_{eff}

Calculation of amplification at medium frequencies:
f = 1000 Hz (ω = 6280 radian/s) and f = 20 Hz (ω = 125.6 radian/s)

Anode current at f = 1000 Hz = (***6280 radian/s*** *) is:*

$$Z_a = R_{DC} + j\omega L; \quad U_g = U_g + j\theta$$

$$i_a=\frac{\mu\times U_g+j\theta}{R_i+\left(R_{DC}+j\omega L\right)}=\frac{41\times 1+j0}{960+\left(700+j6280\times 50\right)}=\frac{41}{1660+j314000}\times\frac{1660-j314000}{1660-j314000}=$$

$$\frac{41\times 1660}{1660^2+314000^2}-j\frac{41\times 314000}{1660^2+314000^2}=\left(0.69\times 10^{-6}-j0.1\times 10^{-3}\right)(A)$$

$$U_{out}=-\left(R_{DC}+j\omega L\right)\times i_a=-\left(700+j6280\times 50\right)\times\left(0.69\times 10^{-3}-j0.13\right)\times 10^{-3}$$

$$U_{out}=40.82-j0.125$$

$$U_{out}=\sqrt{40.82^2+0.125^2}=40.82 \qquad \tan\varphi=\frac{-0.125}{40.82}\rightarrow\varphi=-0.175^0$$

Amplification is:

$$A_{f=1000Hz}=\frac{U_{out}}{U_{in}}=\frac{40.82\angle-0.175^0}{1\angle 0^0}=\mathbf{40.82\angle-0.175^0}$$

Amplification at f = 20Hz is: $A_{f=20\,Hz}=39.87\angle-8.45^0$

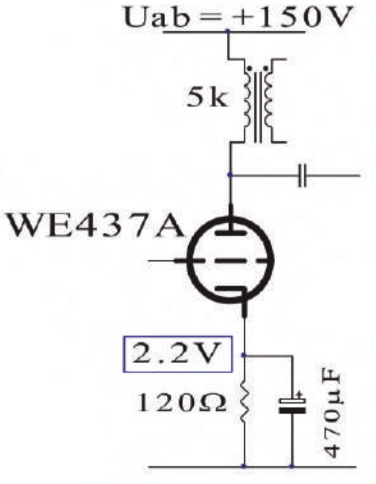
The impedance of the inductance used as the anode load is frequency dependent and therefore the amplification is not equal at different frequencies:

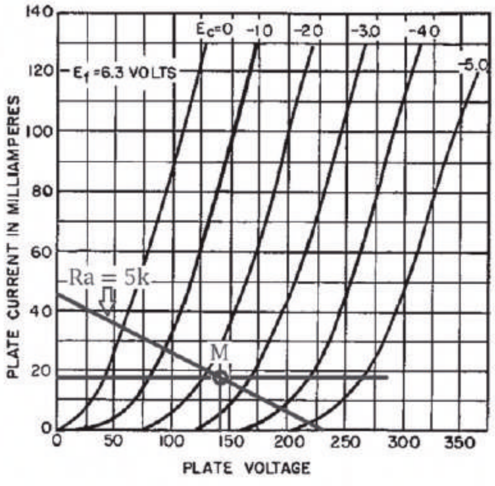
$$\frac{A_{20\,Hz}}{A_{1\,kHz}}=\frac{39.87}{40.82}=0.976=-0.2\,dB$$

(The amplification at 20Hz is 0.2dB lower than the amplification at 1000Hz).

2. Transformer

A transformer as an electronic component used as an anode load is defined by the impedance of the primary (the secondary is loaded with the corresponding load). A transformer used as an anode load can be treated like any other electronic component with impedance $R_a = R_P$ (impedance of the primary in a limited frequency range) so graphical analysis can be applied.

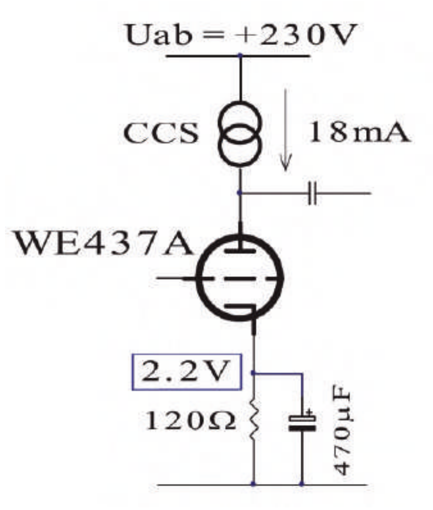


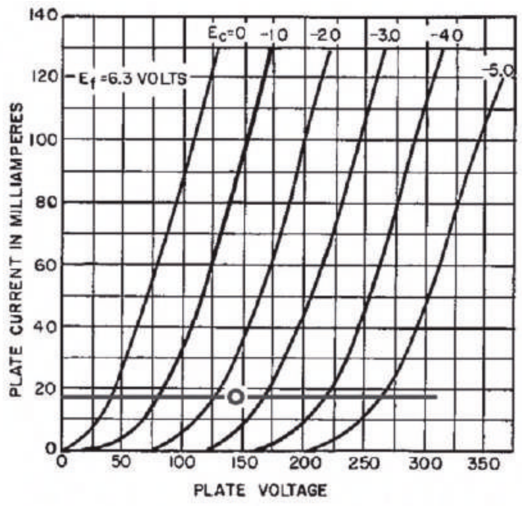


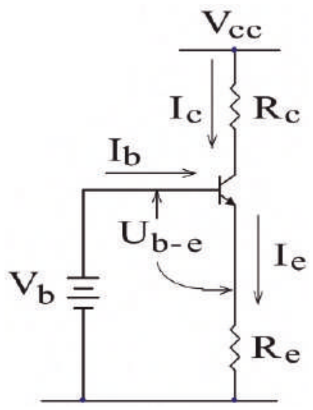
3. Constant Current Source (CCS)

If a constant current source is used as an anode load, theoretically, the load line is a straight horizontal line parallel to the U_a axis on the graph of the anode characteristics. The main feature of commonly used constant current source circuits is their very high impedance.

The use of active loads has several advantages over a resistor used as an anode load: higher amplification *almost equal to the* μ of *the tube, lower harmonic distortion and higher power supply ripple voltage rejection.*







A very simple CCS can be made using a transistor and a few electronic components:

$$V_b - U_{b-e} - R_e \times I_e = 0$$
$$I_e = (h_{fe} + 1) \times I_b$$
$$I_c = h_{fe} \times I_b$$

By solving the above system of equations, I_c (I_c= I_{Load}) is:

$$I_{Load} = I_c = \frac{h_{fe}}{h_{fe} + 1} \times \frac{V_b - U_{b-e}}{R_e}$$

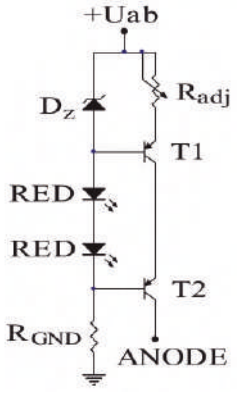
$$I_{Load} = I_c = \frac{h_{fe}}{h_{fe}+1} \times \frac{V_b - 0.6V}{R_e} \approx \frac{V_b - 0.6V}{R_e}$$

By analyzing the above equation, it can be concluded that the current flowing through the collector load is independent of R_L. It can also be concluded: if the voltage at the base of the transistor is constant, the collector current is constant and independent of R_L.

A constant voltage at the base of transistor can be made by using some constant voltage source circuit or by using a Zener diode or some other reference voltage source or by some other electronic components (LED).

CCS, which is commonly used as an anode load, must fulfilled some special technical requirements (like other types of anode load) related to the power dissipation and the voltage across the CCS - one high voltage power transistor should be added to the basic circuit explained above.

CCS design 1:



- $R_{adj} = \frac{V_Z - U_{(b-e)T_1}}{I_{Load}}$

- $R_{GND} = \frac{U_{ab} - V_Z - 2 \times V_{F(LED)}}{I_D}$

I_D *- Current flowing through Zener diode and LEDs: (3.5 ÷ 5) mA*

- Output impedance:

$$R_{out} \approx R_{adj} \times h_{fe(T1)} \times h_{fe(T2)}$$

Example:
Grounded cathode amplifier
Tube:WE 437A
Quiescent point: $U_{a\text{-}k}$ *=140 V,* I_a*=18 mA,* $U_{g\text{-}k}$ *= - 2.2 V*
U_{ab} *= +240 V*

Electronic components:
T1 – BC557B (High gain: $h_{fe(T1)} = (220 \div 475)$*)*
T2 – MJE350 (High voltage: $V_{c\text{-}e} = -300$ *V; High power* $P_c = 20$ *W;* $h_{fe(T2)} = (30 \div 240)$*)*
D_{Z1} *– Zener 5v6, 500 mW*
LED – RED LED 3 mm, U_F *=1.7 V*

$$R_{gnd} = \frac{U_{ab} - V_z - 2 \times V_{F(LED)}}{I_D} = \frac{240 - 5.6 - 2 \times 1.7}{0.004} = 57750\,\Omega$$

Standard: R_{gnd} ***= 62 kΩ***

Dissipation of R_{gnd} ***:***

$$P_{R_{gnd}} = R_{gnd} \times I_D^2 = 62000 \times 0.004^2 = 0.99\,W$$

R_{gnd} ***= 62 kΩ / (4 ÷ 5) W***

$$R_{adj} = \frac{V_Z - U_{(b-e)T1}}{I_{Load}} = \frac{5.6 - 0.6}{0.018} = 277\,\Omega$$

Standard: 500 Ω multi-turn trimmer potentiometer

Calculation of safe dissipation of T2:

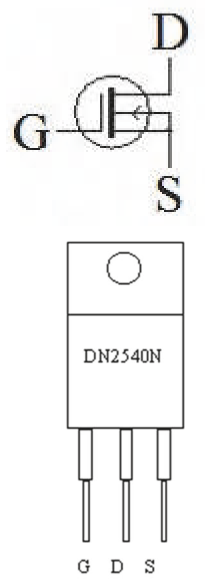
$$P_{d(T2)} = (U_{ab} - V_Z - 2 \times V_F - U_{a-k}) \times I_{Load} = (240 - 5.6 - 3.4 - 140) \times 0.018 = 1.64\,W$$

 # T2 should be placed on the appropriate heat sink.

$$R_{out(min)} \approx R_{adj} \times h_{fe(T1)} \times h_{fe(T2)} = 277 \times 220 \times 30 = 1.8\,M\Omega$$

CCS design 2:
N – Channel Depletion – Mode Vertical DMOS FET can be used to design CCS. To avoid the problems caused by the high shunt capacitance of the DMOS FET, a cascade configuration circuit should be used.

One of the most popular DMOS FET used in CCS applications is: DN2540.

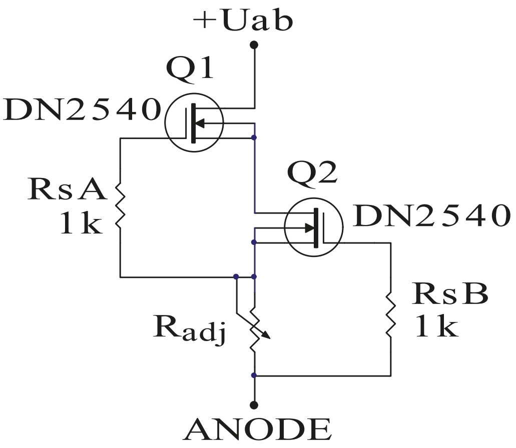


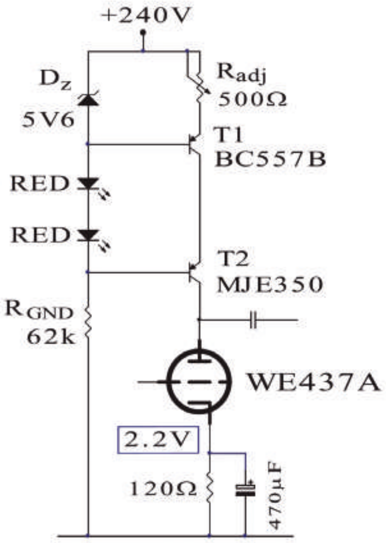
$$R_{adj} \approx \frac{2V}{I_{Load}}$$

Example above:

$$R_{adj} \approx \frac{2V}{I_{Load}} \approx \frac{2}{0.018} \approx \mathbf{111\ \Omega}$$

Standard: 250 Ω, mult-turn trim. pot.

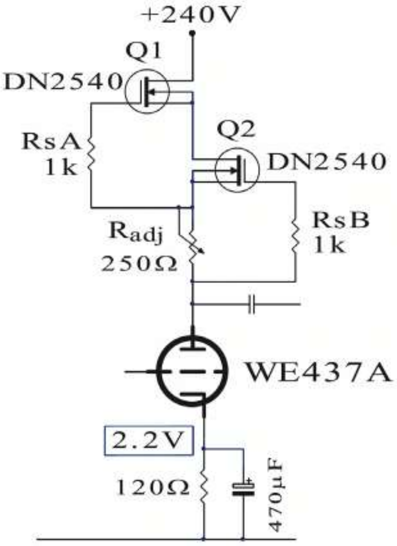


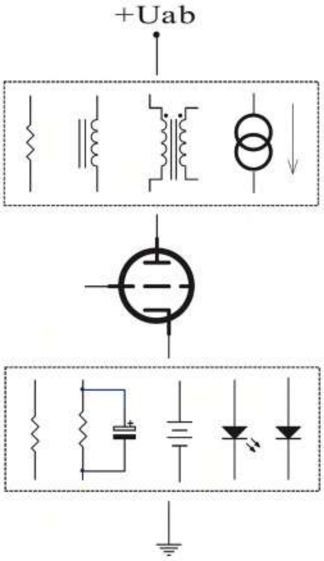


Conclusion:

The flexibility of the grounded cathode (anode follower) amplifier configuration gives the designer a lot of freedom to make the amplifier the way he wants it. Which type of anode load and bias circuit the designer chooses depends on the skills and technical knowledge of the designer and specific technical requirements and characteristics of the amplifier that must be fulfilled. Anyway, the only advice to designers is:

keep your mind open to any new idea, dream idea or idea from your imagination.





4.2 GROUNDED-ANODE (cathode follower)

BASIC CIRCUIT

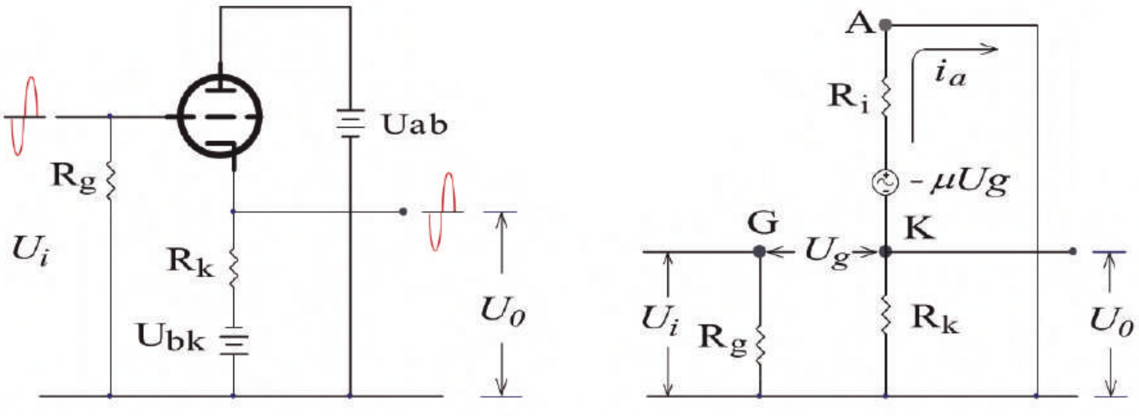


Fig. 4-10

Analyzing the circuit shown in Fig.4-10, the anode current is: $i_a = \frac{-\mu \times U_g}{R_i + R_k}$. Also, the grid – cathode voltage is equal to the input voltage plus the voltage across the cathode resistor $U_g = U_i + i_a \times R_k$, so:

$$i_a = \frac{-\mu \times U_i}{(1+\mu) \times R_k + R_i} \longrightarrow U_i = \frac{-i_a \times [(1+\mu) \times R_k + R_i]}{\mu}$$

The output voltage caused by the flow of current i_a through the resistor R_k equals: $U_0 = \frac{-\mu \times U_i \times R_k}{(1+\mu) \times R_k + R_i}$

Amplification is:

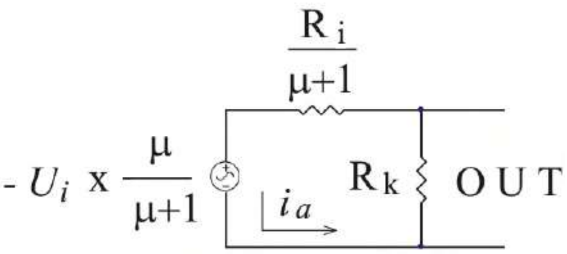
$$A = \frac{U_0}{U_i} = \frac{\mu}{(1+\mu) + \frac{R_i}{R_k}} \qquad (401.21)$$

The amplification is less than 1 and the output voltage is **in phase** with the input voltage.

 # Note: This circuit is also called a ***Cathode follower*** *because the cathode voltage follows the grid voltage.*

The calculation of the output impedance of the grounded anode amplifier can be simplified by redesigning the equivalent circuit.

The equation $i_a = \frac{-\mu \times U_i}{(1+\mu) \times R_k + R_i}$ *can be written as* $i_a = \frac{\frac{-\mu \times U_i}{\mu + 1}}{R_k + \frac{R_i}{(1+\mu)}} = \frac{U_{gen}}{R_{gen}}$ *and the electric diagram of the redesigned equivalent circuit looks like this:*



By analyzing the redesigned equivalent circuit it can be concluded that the output impedance is equivalent to the parallel connection R_k and $\frac{R_i}{\mu+1}$:

$$R_{out}=R_k \parallel \frac{R_i}{\mu+1}=\frac{R_k\times\frac{R_i}{\mu+1}}{R_k+\frac{R_i}{\mu+1}}=\frac{R_k\times R_i}{(\mu+1)\times R_k+R_i}$$

$$R_{out}=\frac{R_i}{(\mu+1)+\frac{R_i}{R_k}} \qquad (401.22)$$

The output impedance is very low (minimum μ+1 times lower than the internal resistance R_i of the tube).

The grounded anode amplifier, because of its characteristics: high input and very low output impedance, can be used as an impedance matching circuit.

The **input impedance** of this circuit is high, the **voltage amplification** is **almost equal to 1** and the **output impedance** is **low**. It acts as a low impedance signal source.

The circuit is used as an impedance matching circuit, to connect a high output impedance signal source to a low input impedance circuit.

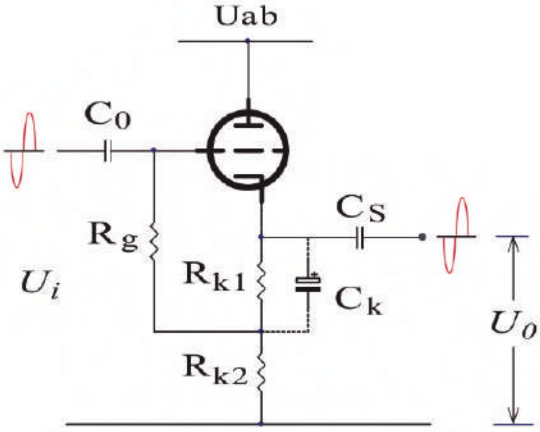


Fig. 4-11

In practical applications, a redesigned base circuit is usually used (to avoid the use of an additional voltage source in the cathode circuit).

Grid bias is determined by R_k1 and quiescent current:

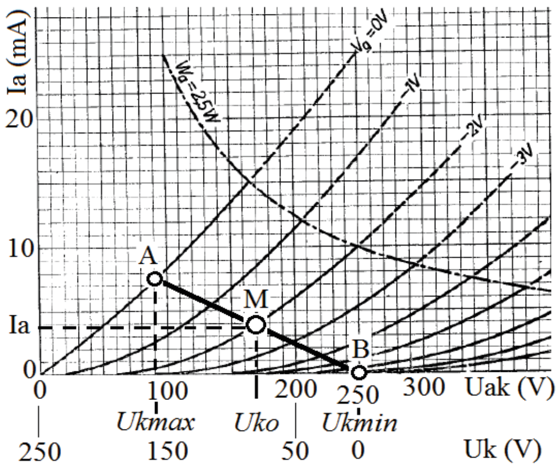
$$-U_{g-k}=R_k1\times I_a$$

So the graphical analysis is similar to the graphical analysis of a ground cathode amplifier.

A positive feature of this circuit is its high input impedance:

$$R_{in}=(\mu+1)\times(R_g+R_{k2})$$

Graphical analysis



In the characteristics diagram $I_a = f(U_a)$, the abscissa is the **voltage between the anode and the cathode** instead of the commonly used anode voltage. DC cathode voltage shown on the abscissa can be calculated as:

$$U_k = U_{ab} - U_{ak} \tag{401.23}$$

The input AC voltage **$U_{in} = U_g$** is equal to the sum of the output AC voltage **$U_{out} = U_k$** and the AC voltage between the grid and the cathode **U_{gk}:**

$$U_g = U_{out} + U_{gk} \tag{401.24}$$

If the cathode DC voltage is denoted as **U_{k0},** the DC voltage between the grid and the cathode (grid bias voltage) as **U_{gk0}**, the instantaneous voltage at the cathode as u_k and the instantaneous voltage between the grid and the cathode as u_{gk}:

$$U_k = u_k - U_{k0} \text{ and } U_{gk} = u_{gk} - U_{gk0} \text{ .}$$

By substituting U_{out} and U_{gk} of the equation (401.24) with the previous equations:

$$U_g = u_k - U_{k0} + u_{gk} - U_{gk0} \tag{401.25}$$

The operation of the tube in class A is limited by $u_{gk} = 0$ and $u_{gk} = u_{gk\,min}$ ($u_{gk\,min}$: near I_a cutoff), so:

$$U_g = u_{kmax} - U_{k0} + 0 - U_{gk0}$$
$$- U_g = u_{kmin} - U_{k0} + u_{gkmin} - U_{gk0}$$

By summing the above two equations:

$$U_{k0} = \frac{U_{k\,max} + U_{k\,min}}{2} - \left(U_{gk\,0} - \frac{U_{gk\,min}}{2}\right) \approx \frac{U_{k\,max} + U_{k\,min}}{2} \tag{401.26}$$

The slope of the load line is determined by **R_k**. ($R_k = R_{k1} + R_{k2}$); R_{k1} is obtained from:

$$R_{k1} = \frac{U_{gk0}}{I_a} \tag{401.27}$$

Using the graphic above:

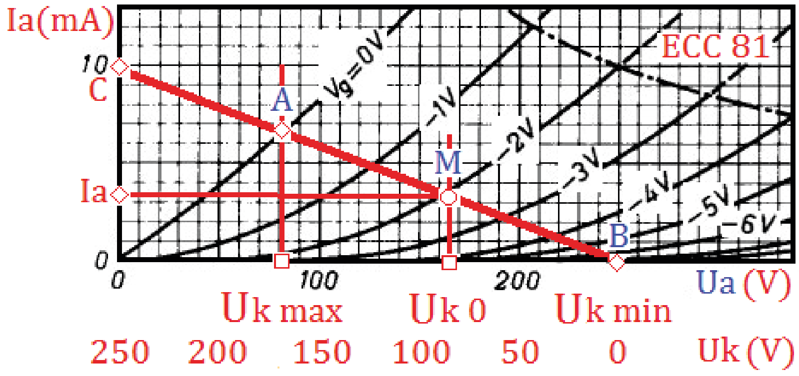
u_{kmin} = 6 V; u_{kmax} = 161 V; U_{gk0} = - 2 V

$$U_{k0} = \frac{U_{k\,max} + U_{k\,min}}{2} = \frac{161 + 6}{2} = 83.5\,V$$

R_k = 22.6 kΩ (slope of the load line); $R_{k1} = \frac{U_{gk0}}{I_a} = \frac{2\,V}{3.7\,mA} = 588\,\Omega$

 # Note:
The first impression after reading the above text is that the process of graphical analysis of grounded anode amplifier is very complicated, but it is not rally so. Its application in the practical process of designing a conventional grounded anode amplifier is not so complicated.

Example 402.1:
Grounded anode amplifier (Fig.4-11); Design requirement: $U_k = 85$ V .
Tube: 1 / 2 ECC81 (12AT7), (one triode).
$\mu = 60$, $R_i = 16$ kΩ
Available power supply: $U_{ab} = 250$ V



1. The voltage U_{ak} (voltage between anode and cathode of the tube (triode)) is approximately equal:

   $U_{ab} - U_{k0} = 250 - 85 = 165$ V

2. The procedure for determining and drawing the load line is similar to the procedure used in the case of a grounded cathode amplifier. In the characteristics diagram $I_a = f(U_a)$, the abscissa is the voltage between the anode and the cathode.

   The cathode DC voltage is: $U_k = U_{ab} - U_{ak}$
   The point **B** equal to the cathode voltage $U_k = 0$V is set at the same place where the point U_{ab} is set on the abscissa of the original characteristics diagram $I_a = f(U_a)$, (DC power supply voltage).

3. Draw a vertical line passing through the point of abscissa $U_{k0} = 85$ V.

4. Using equation (401.26):

$$U_{k0}=\frac{U_{k\,max}+U_{k\,min}}{2}-\left(U_{gk0}-\frac{U_{gk\,min}}{2}\right)\approx\frac{U_{k\,max}+U_{k\,min}}{2}\rightarrow U_{kmax}=2\times U_{k0}-U_{k\,min}$$

or $U_{kmax}\approx 2\times U_{k0}$

Draw a vertical line passing through the point of abscissa $U_{k\,max} = 2 \times U_{k0} = 2 \times 85\text{ V} = 170$ V.
This vertical line intersects the anode characteristic $U_g = 0$V at point **A**.

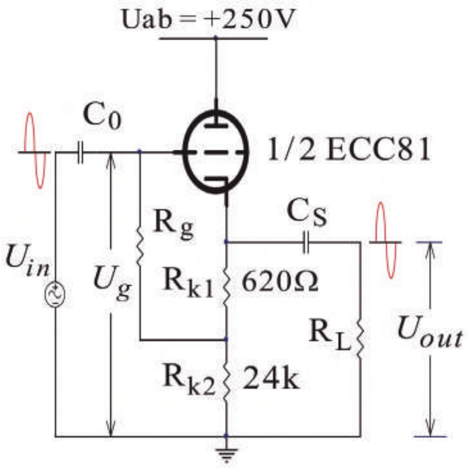
Draw a straight line passing through points **B** [$U_k = 0$V ($U_{ab} = 250$V)] and **A** – this is the **load (working) line.**
The load line intersects the I_a axis at point **C**: $I_a =$ **9.9 mA**. The slope of the load line is determined by the cathode resistor $\mathbf{R_k}$, so:

$$R_k=\frac{U_{ab}}{9.9\,mA}=\frac{250\,V}{9.9\,mA}=25\,k\Omega$$

5. The vertical line passing through the point $\mathbf{U_{k0} = 85V}$ intersects the load line at the point **M** located on the anode characteristic $\mathbf{U_g = -2.1\ V}$, i.e. **M (85 V, 3. 3mA)**.

6. Calculation of $\mathbf{R_{k1}}$: $R_{k1} = U_{gk} / I_a = 2.1 / 3.3 = 0.636$ kΩ (standard: $\mathbf{R_{k1} = 620}$ Ω)

   $R_k = R_{k1} + R_{k2} \rightarrow R_{k2} = R_k - R_{k1} = 25\text{ k}\Omega - 0.62\text{ k}\Omega = 24.38$ kΩ; Standard: $\mathbf{R_{k2} = 24\ k}$



7. Calculation of amplification (**A**):

$$A=\frac{\mu}{(1+\mu)+\frac{R_i}{R_k}}=\frac{60}{(60+1)+\frac{16k}{24k}}=0.973$$

8. Calculation of **R$_{out}$** :

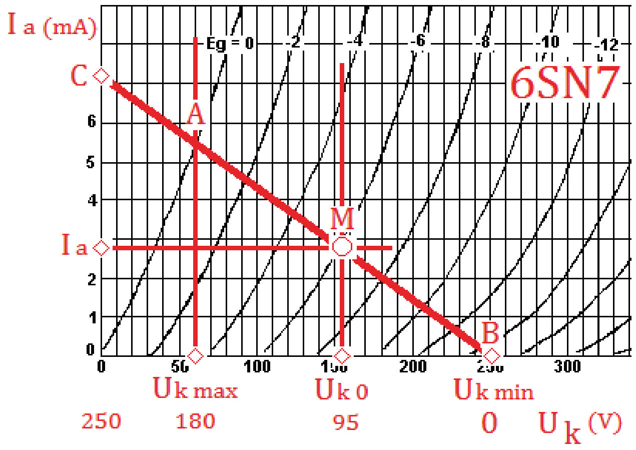
$$R_{out}=\frac{\frac{R_i}{\mu+1}\times R_k}{\frac{R_i}{\mu+1}+R_k}=\frac{\frac{16k}{60+1}\times 24k}{\frac{16k}{60+1}+24k}=0.259\,k\Omega$$

Example 402.2:

Grounded anode amplifier (Fig.4-10); $U_k = 95$ V .

Tube: 1 / 2 6SN7 (one triode); $R_i = 7.7$ kΩ; μ = 20

Available power supply: $U_{ab} = 250$ V



1. Anode - cathode voltage of the tube is approximately:
   $U_{ab} - U_{k0} = 250 - 95 = 155$ V
2. The procedure for determining and drawing the load line is similar to the procedure used in the case of grounded cathode amplifier.
   In the diagram of the anode characteristics $I_a = f(U_a)$, the abscissa is the voltage between the anode and the cathode.
   The cathode DC voltage is: $U_k = U_{ab} - U_{ak}$
   The point **B** equal to the cathode voltage $U_k = 0$V is set at the same place where the point U_{ab} is set on the abscissa.
3. Draw a vertical line passing through the point of abscissa $U_{k0} = 95$ V.
4. Using equation (401.26):

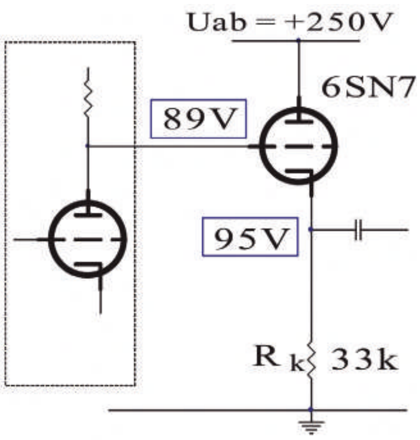
$$U_{k0}=\frac{U_{k\,max}+U_{k\,min}}{2}-\left(U_{gk0}-\frac{U_{gk\,min}}{2}\right)\approx\frac{U_{k\,max}+U_{k\,min}}{2}\rightarrow U_{kmax}=2\times U_{k0}-U_{k\,min}$$

or $U_{kmax}\approx 2\times U_{k0}$

Draw a vertical line passing through the point of abscissa $U_{k\,max} = 2 \times U_{k0} = 2 \times 95\text{ V} = 180$ V.
This vertical line intersects the anode characteristic $U_g = 0$V at point **A**.
Draw a straight line passing through points **B** [$U_k = 0$V ($U_{ab} = 250$V)] and **A** – this is the **load line**.
The load line intersects the I_a axis at point **C**: $I_a =$ **7.2 mA**. The slope of the load line is determined by the cathode resistor **R$_k$** , so:

$$R_k=\frac{U_{ab}}{7.2\,mA}=\frac{250\,V}{7.2\,mA}=34.7\,k\Omega$$

Standard: $R_k = 33$ kΩ



5. The vertical line passing through the point **U_{k0} = 95 V** intersects the load line at the point **M** located on the anode characteristic **U_g = - 6.1 V**, i.e. **M (95 V, 2.8 mA)**.

6. Calculation of amplification (**A**):

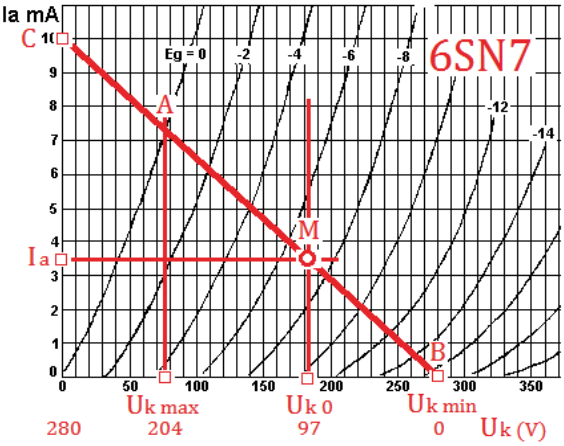
$$A=\frac{U_0}{U_i}=\frac{\mu}{(1+\mu)+\frac{R_i}{R_k}}=\frac{20}{(20+1)+\frac{7.7k}{33k}}=0.94$$

7. Calculation of **R_{out}** :

$$R_{out}=\frac{\frac{R_i}{\mu+1}\times R_k}{\frac{R_i}{\mu+1}+R_k}=\frac{\frac{7.7k}{20+1}\times 33k}{\frac{7.7k}{20+1}+33k}=0.36\,k\Omega$$

Example 402.3:

Grounded anode amplifier (Fig.4-11)
Design requirements: U_{in}= 5V_{RMS} and U_k = 97V.
Tube: 1 / 2 6SN7 (one triode).
R_i = 7.7 kΩ, μ = 20
Available power supply: U_{ab} = 280V



1. Calculation of U_{gk} that fulfills the first design requirement U_{in} = **5 V_{RMS}** :

$$U_{gk} = \sqrt{2} \times U_{in(RMS)} = 1.41 \times 5 = 7V$$

This means that the working point (quiescent point) must be located somewhere on the anode characteristic U_g = -7 V.

2. Draw a vertical line passing through the point of abscissa **U_{k0} = 97V** to fulfill the second design condition **U_k = 97V**.

3. The vertical line passing through the point **U_{k0} = 97V** intersects the anode characteristic U_g = -7V at the point M (97V, 3.5mA) – quiescent point.

4. Draw a straight line passing through points **B (U_k = 0V)** of the abscissa and **M (97V, 3.5mA)**. This is the **load line** and it intersects the I_a axis at the point **C**: $I_{a(C))}$ = 10mA. The slope of the load line is determined by the sum of cathode resistors (**R_{k1} + R_{k2}**), so:

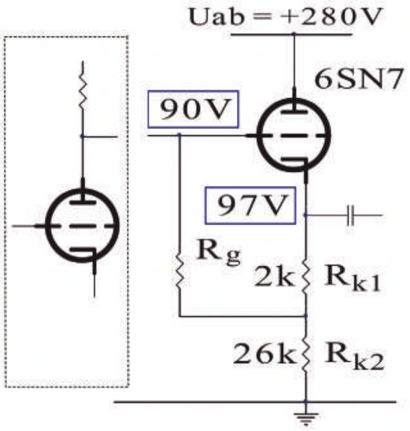
$$R_{k1}+R_{k2}=\frac{U_{ab}}{I_{a(C)}}=\frac{280\,V}{10\,mA}=28\,k\Omega$$

5. Calculation of **R_{k1} :**

$$R_{k1}=\frac{U_g}{I_a}=\frac{7V}{3.5\,mA}=2\,k\Omega$$

6. Calculation of **R_{k2} :**

$$R_{k2}=28\,k\Omega-R_{k1}=28\,k\Omega-2\,k\Omega=26\,k\Omega$$



7. Calculation of amplification **A**:

$$A=\frac{\mu}{(1+\mu)+\frac{R_i}{R_k}}=\frac{20}{(20+1)+\frac{7.7k}{28k}}=0.94$$

8. Calculation of **R_{out}** :

$$R_{out}=\frac{\frac{R_i}{\mu+1}\times R_k}{\frac{R_i}{\mu+1}+R_k}=\frac{\frac{7.7k}{20+1}\times 28k}{\frac{7.7k}{20+1}+28k}=0.362\,k\Omega$$

Note:
When a DC voltage source is used to supply the tube heater, the condition relating to the maximum permissible voltage between the heater and the cathode - V_{kf} (published in the tube data sheet as the limit voltage V_{kf}) must be fulfilled.

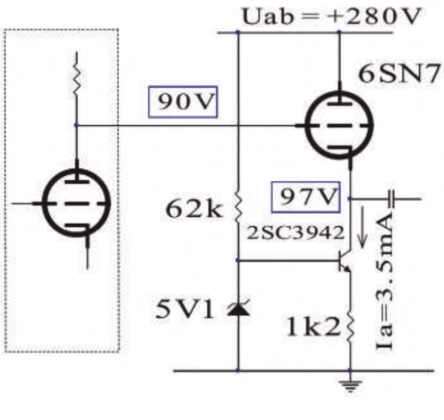
The voltage between the heater and the cathode must be lower than max. U_{kf} (limit voltage U_{kf}).
In order to fulfill the above condition, the supply voltage of the heater must be virtually elevated to an electrical potential that ensures that the difference between the electrical potentials of the cathode and the heater is lower than max. V_{kf} .

The cathode resistor can be replaced by a constant current source (CCS).

Example 402.4
Grounded anode amplifier (Fig.4-10).
Design requirements: U_k = 97V
(the designed amplifier should be DC coupled to the previous amplifier stage).
Tube: 1 / 2 6SN7, (one triode).
Available power supply: U_{ab} = 280V

Previous example (402.3): 6SN7 operates at U_{ak} = 183V and I_a = 3.5mA at U_{gk} = -7V.

The same above operating conditions of the 6SN7 can be achieved by making a constant current source (I = 3.5mA) and inserting it into the cathode circuit instead of a resistor.



Note:
The U_{ce} of the transistor used in CCS must be higher than the designed U_k .
Any well-filtered and stable DC voltage source (eg heater DC power supply) can be used as a constant voltage source in the base circuit of the transistor.

4.3 GROUNDED-GRID

BASIC CIRCUIT

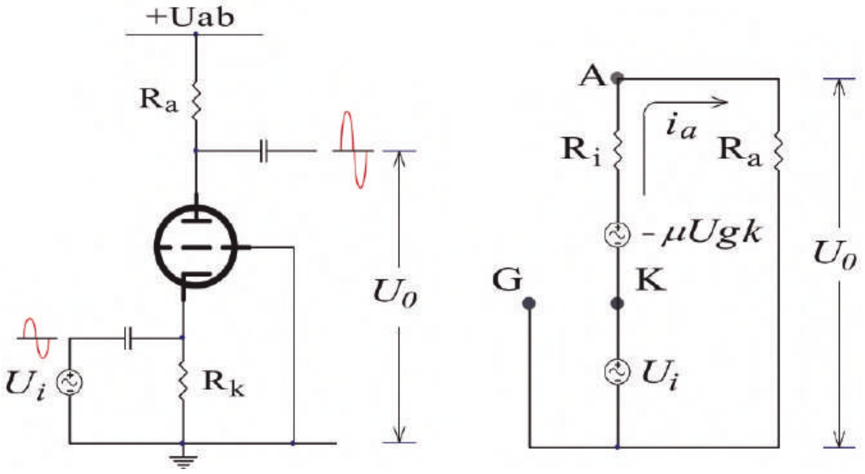


Fig. 4-12

Equivalent circuit: $U_{gk} = U_i$ and $U_i - \mu \times U_{gk} - I_a \times R_i - I_a \times R_a = 0$ i.e.:

$$U_i - \mu \times U_{in} - I_a \times R_i - I_a \times R_a = 0 \qquad (401.28)$$

The output voltage is: $U_0 = I_a \times R_a$. Combining this equation with the previous one:

$$U_i \times (\mu+1) - \frac{(R_i+R_a)}{R_a} \times U_0 = 0$$

Amplification is:

$$A = \frac{U_0}{U_i} = (\mu+1) \times \frac{R_a}{R_i+R_a} \qquad (401.29)$$

Input impedance:

By solving equation 401.28 for $\frac{U_i}{I_a}$:

$$R_{in} = \frac{U_i}{I_a} = \frac{R_i+R_a}{\mu+1} : \qquad (401.30)$$

Main characteristics:

- **Input and output signals are in phase.**
- The input impedance is low and the output impedance is high.

A grounded grid amplifier can be used in application where the impedance of the input source is very low, as a stand-alone, but a more useful application of this circuit is in combination with some other types of amplifier circuits (combined with a common cathode amplifier circuit, for example).

 # Note1:
Optimal characteristics are achieved by using tubes with high μ and low internal resistance R_i (WE417A, 5842, 6C45П, EC900, ECC88…)

 # Note2:
In order to avoid problems at low input frequencies caused by low R_{in} , it is necessary to use a high capacitance coupling capacitor at the input of the circuit (of the order of tens to hundreds μF).

Example 403.1:
Low input impedance amplifier circuit (grounded grid):
A ≈ 37, $R_{in} \geq$ 60 Ω and U_{in} = 700 mV_{RMS}
Tube: 5842 (WE417A); [Ri = 1.7 kΩ, μ = 43]

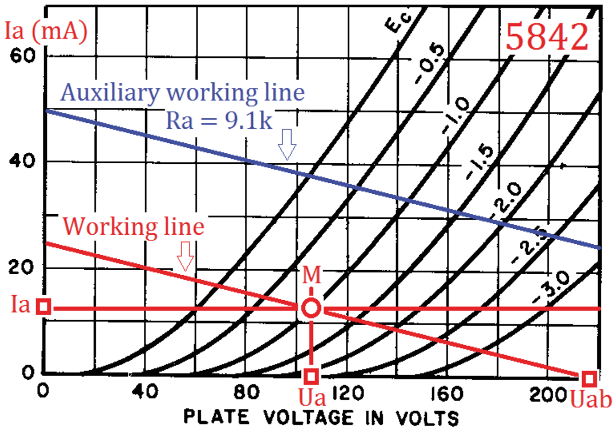
1. Using the amplification equation:

$$A=\frac{(\mu+1)\times R_a}{R_i+R_a} \rightarrow R_a=\frac{A\times R_i}{(\mu+1)-A}=\frac{37\times 1.7k}{(43+1)-37}=9k$$ (Standard: Ra = 9.1 kΩ)

2. Using the R_{in} equation:

$$R_{in}=\frac{R_k\times\frac{R_i+R_a}{\mu+1}}{R_k+\frac{R_i+R_a}{\mu+1}} \rightarrow \frac{R_k\times\frac{1.7k+9.1k}{43+1}}{R_k+\frac{1.7k+9.1k}{43+1}}\geq 0.06\ k\Omega \rightarrow R_k\geq 79.5\Omega$$ (Standard: R_k = 82 Ω)

3. Draw an "**auxiliary**" load line with a "**slope**" of **9.1 kΩ** somewhere on the Ia = f(U_a) characteristic.



4. The condition of holding an input signal of 700 mV_{RMS} is fulfilled if:

$$U_{gk}\geq U_{in\ Peak}=\sqrt{2}\times U_{in\ max}$$

$$=1.41\times 0.=0.99V \quad \rightarrow$$

$$I_a=\frac{U_{gk}}{R_k}=\frac{0.99V}{80\Omega}=12mA$$

5. Draw a line $\mathbf{I_a}$ **= 12mA** parallel to the U_a axis: this line intersects I_a = f(U_a) characteristic $\mathbf{U_g}$ **= - 0.99V** at the point: **M (105V, 12mA)**.

6. Draw a line parallel to the "auxiliary" load line passing through point M – this is the **load line**. The load line intersects the U_a axis at $\mathbf{U_{ab}}$ **≈ 215V**. So, the required power supply is 215V.

7. Recalculate A and R_{in}:

$$A=\frac{(\mu+1)\times R_a}{R_i+R_a}=\frac{(43+1)\times 9.1k}{1.7k+9.1k}=37.07$$

$$R_{in}=\frac{R_k\times\frac{R_i+R_a}{\mu+1}}{R_k+\frac{R_i+R_a}{\mu+1}}=\frac{0.082k\times\frac{1.7k+9.1k}{43+1}}{0.082\ \ \frac{1.7k+9.1k}{43+1}}=0.061\ k\Omega$$

 # Note:
If a tube with low internal resistance and high μ is not available, the standard ECC 83 (12AX7) may be a good choice as will be shown in the following example.

Example 403.2:
Low input impedance amplifier circuit:
A ≈ 60, $R_{in} \geq$ 300Ω and U_{in} = 350 mV_{RMS}
Tube: ECC83

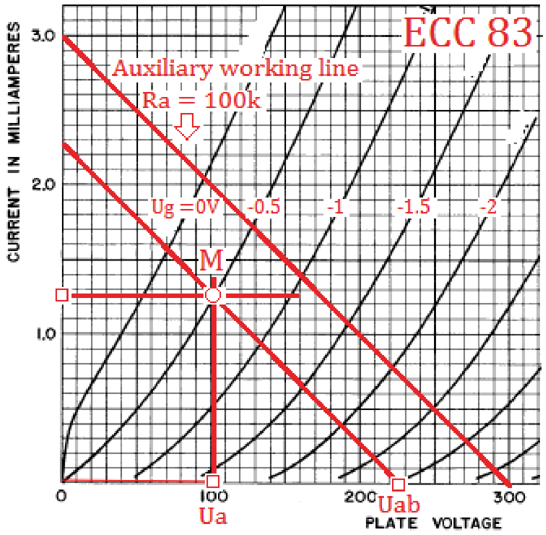
1. Using the amplification equation:

$$A=\frac{(\mu+1)\times R_a}{R_i+R_a}\rightarrow R_a=\frac{A\times R_i}{(\mu+1)-A}=\frac{60\times 62.5\,k}{(100+1)-60}=91.46\,k\Omega$$ (Standard: R_a = 100 kΩ)

2. Using the R_{in} equation:

$$R_{in}=\frac{R_k\times\frac{R_i+R_a}{\mu+1}}{R_k+\frac{R_i+R_a}{\mu+1}}\rightarrow\frac{R_k\times\frac{62.5k+100k}{100+1}}{R_k+\frac{62.5k+100k}{100+1}}\geq 0.3\ k\Omega\rightarrow R_k\geq 369\ \Omega$$ (Standard: R_k = 390 Ω)

3. Draw an “**auxiliary**” load line with a “**slope**” of **100 kΩ** somewhere on the $I_a = f(U_a)$ characteristic.



4. The condition of holding the input signal of 350mV$_{RMS}$ is fulfilled if:

$$U_{gk}\geq U_{in\ Peak}=\sqrt{2}\times U_{in\ max}$$

$$=1.41\times 0.35=0.49V\ \rightarrow$$

$$I_a=\frac{U_{gk}}{R_k}=\frac{0.49V}{390\Omega}=1.26mA$$

5. Draw a line **I_a = 1.26mA** parallel to the U_a axis. This line intersects $I_a = f(U_a)$ characteristic **U_g = - 0.49V** at the point: **M (100V, 1.26mA)**.

6. Draw a line parallel to the “auxiliary” load line passing through point M – this is the **load line**. The load line intersects the U_a axis at **$U_{ab} \approx$ 225V**. So, the required power supply is 225V.

7. Recalculate A and R_{in}:

$$A=\frac{(\mu+1)\times R_a}{R_i+R_a}=\frac{(100+1)\times 100k}{62.5k+100k}=62$$

$$R_{in}=\frac{R_k\times\frac{R_i+R_a}{\mu+1}}{R_k+\frac{R_i+R_a}{\mu+1}}=\frac{0.39k\times\frac{62.5k+100k}{100+1}}{0.39k+\frac{62.5k+100k}{100+1}}=0.314\ k\Omega$$

The above circuit (If R_k is determined depending on the impedance of the MC cartridge used) can be used as a moving coil preamplifier (the next amplifier stage should be RIAA equalized, of course).

4.4 CASCODE

BASIC CIRCUIT

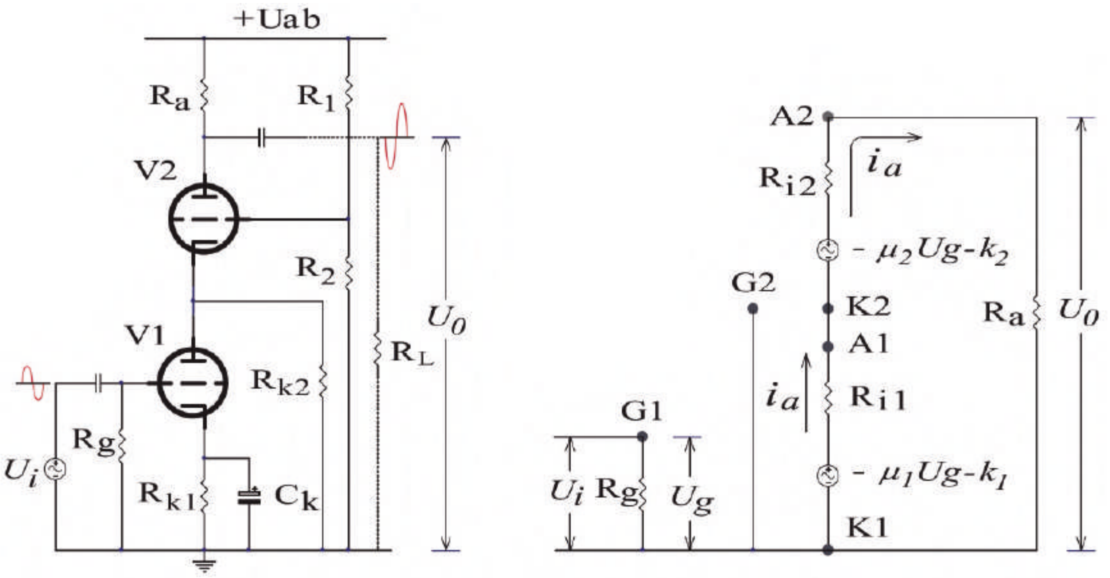


Fig.4-13

The cascode amplifier is a combination of two amplifier circuits already analyzed:

- the lower triode is connected as a grounded cathode amplifier
- the upper triode is connected as a grounded grid amplifier (acts as the anode load of the lower tube).

The anode load (R_{a1}) of the lower triode (V1) is a parallel connection of the input impedance of the grounded grid amplifier circuit (V2) and R_{k2}:

$$R_{a1}=\frac{R_{k2}\times\frac{R_{i2}+R_a}{\mu_2+1}}{R_{k2}+\frac{R_{i2}+R_a}{\mu_2+1}} \tag{401.31}$$

Amplification of the grounded cathode amplifier circuit (V1), A_1:

$$A_{10}=-\mu_1\times\frac{R_{a1}}{R_{i1}+R_{a1}} \tag{401.32}$$

Amplification of the grounded cathode amplifier circuit (V1) without C_k:

$$A_{1r}=\frac{A_{10}}{1+\frac{R_{k1}}{R_{a1}}\times A_{10}} \tag{401.33}$$

Amplification of the grounded grid amplifier circuit (V2), A_2 :

$$A_2=\frac{(\mu_2+1)\times R_a}{R_{i2}+R_a} \tag{401.34}$$

The **total** amplification of the cascode amplifier is:

$$\mathbf{A} = U_{out} / U_{in} = \mathbf{A_{10} \times A_2} \tag{401.35}$$

$$A=-\mu_1\times\frac{R_{a1}}{R_{i1}+R_{a1}}\times\frac{(\mu_2+1)\times R_a}{R_{i2}+R_a}\approx\frac{-\mu_1\times(\mu_2+1)\times R_a}{R_{i2}\times(\mu_2+1)+R_{i2}+R_a} \tag{401.36}$$

If equivalent lower and upper tubes are used, i.e. $\mu_1 = \mu_2 = \mu$ and $R_{i1} = R_{i2} = R_i$:

$$A \approx \frac{-\mu \times (\mu+1) \times R_a}{(\mu+2) \times R_i + R_a} \tag{401.37}$$

or the circuit without C_k:

$$\mathbf{A} = U_{out} / U_{in} = \mathbf{A_{1r} \times A_2} \tag{401.38}$$

The input and output signals are opposite in phase.

***Main application: wideband low noise amplifier.**

Example 404.1:

The circuit Fig.4-14 ; $U_{in} = 0.7\ V_{RMS}$.
Calculate R'_a, R_{a1}, A_{10}, A_{1r}, A_2 and A with and without C_k,

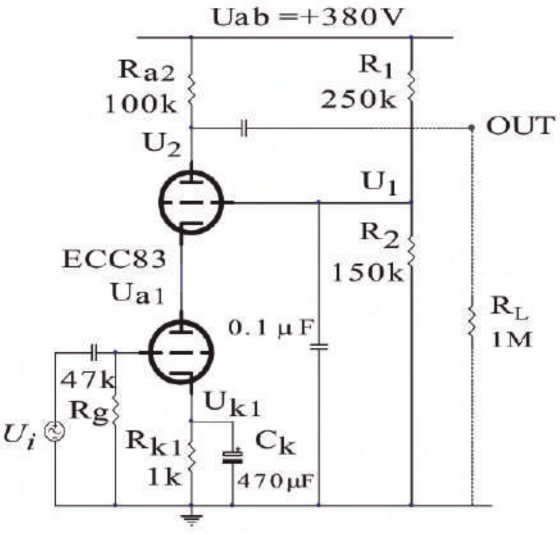


Fig. 4-14

1. $R_a^{'} = R_a \, II \, R_L = \frac{R_a \times R_L}{R_a + R_L} = \frac{100k \times 1000k}{100k + 1000k} = 90.91\,k\Omega$

2. $R_{a1} \approx \frac{R_{i2} + R_a^{'}}{\mu_2 + 1} = \frac{62.5k + 90.91k}{100+1} = 1.52\,k\Omega$

3. $A_{10} = -\mu_1 \times \frac{R_{a1}}{R_{i1} + R_{a1}} = -100 \times \frac{1.52k}{62.5k + 1.52k} = 2.37$

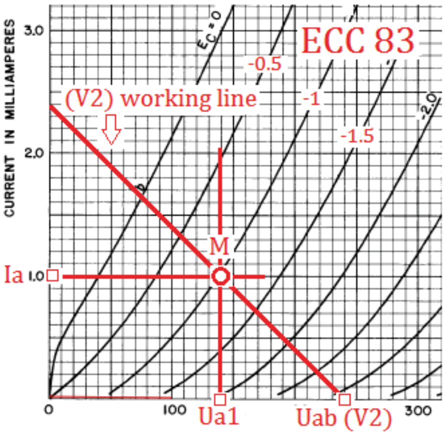
4. $A_{1r} = \frac{A_{10}}{1 + \frac{R_{k1}}{R_{a1}} \times A_{10}} = \frac{2.37}{1 + \frac{1k}{1.52k} \times 2.37} = 0.92$

5. $A_2 = \frac{(\mu_2 + 1) \times R_a^{'}}{R_{i2} + R_a^{'}} = \frac{(100+1) \times 90.91k}{62.5k + 90.91k} = 59.85$

6. $A = A_{10} \times A_2 = 2.37 \times 59.85 = 141.85$

$A_r = A_{1r} \times A_2 = 0.92 \times 59.85 = 55$

7. $U_1 = \frac{R_2}{R_1 + R_2} \times U_{ab} = \frac{150k}{250k + 150k} \times 380\,V = 142.5\,V$



8. Using the diagram $I_a = f(U_a)$ (ECC83 - triode V_1) it can be seen that if the $U_{a1} \approx 140$ V and $U_g \approx -1$ V → $I_a \approx 1$ mA.
   An equal current flows through triode V_2.

   Operating conditions of the second triode (V_2):
   $U_{ab(V2)} \approx U_{ab} - U_1 = 380 - 140 = 240$ V; $R_{a2} = 100$ kΩ;
   $I_a \approx 1$ mA.

   By drawing the load line it can be seen that:
   $U_2 \approx U_1 + 140\ V = 280\ V$.

Example 404.2:
Tube ECC88 (6DJ8) is special quality double triode designed for use as a cascade amplifier circuit (primarily in RF circuits).

High amplification (A ≈ 100) highly sensitive cascode amplifier circuit.

$R_L = 1M\Omega$
Tube: ECC88.
$\mu = 33$
$S = 12.5$ mA/V
* $\mu = S \times R_i \rightarrow R_i = \mu / S$
$R_i = 33 / 12.5 = 2.64$ kΩ

1. Using the amplification equation:

$$A \approx \frac{-\mu \times (\mu+1) \times R_a'}{(\mu+2) \times R_i + R_a'} \rightarrow R_a' = \frac{(\mu+2) \times R_i}{\frac{\mu \times (\mu+1)}{A} - 1} = \frac{(33+2) \times 2.64\,k}{\frac{33 \times (33+1)}{100} - 1} = 9.04\,k\Omega$$

$$R_a' = \frac{R_a \times R_L}{R_a + R_L} \rightarrow R_a = \frac{R_a' \times R_L}{R_L - R_a'} = \frac{9.04\,k \times 1000\,k}{1000\,k - 9.04\,k} = 9.12\,k\Omega$$ (Standard: $R_a = 9.1$ kΩ, or $R_a = 10$ kΩ)

2. Amplification of the upper triode circuit (grounded grid amplifier):

$$A_2 = \frac{(\mu+1) \times R_a'}{R_i + R_a'} = \frac{(33+1) \times 9.1\,k}{2.64\,k + 9.1\,k} = 26.35$$

3. Using the equation for the total amplification of a cascode amplifier:

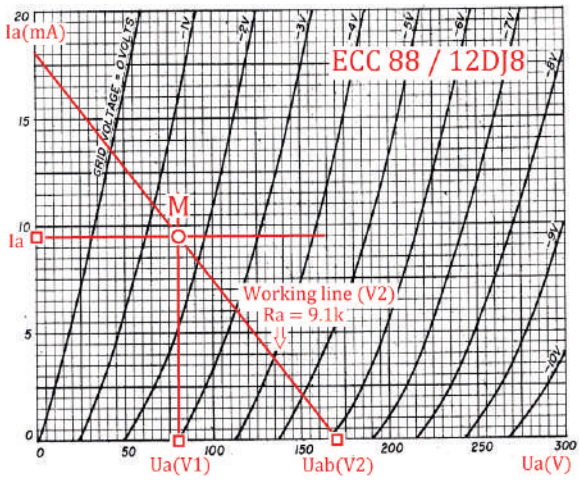
$$A = A_1 \times A_2 \rightarrow A_1 = \frac{A}{A_2} = \frac{100}{26.35} = 3.79$$

4. Using the A_1 equation:

$$A_1 = -\mu_1 \times \frac{R_{a1}}{R_{i1} + R_{a1}} \rightarrow R_{a1} = \frac{A_1 \times R_i}{\mu - A_1} = \frac{3.79\,k \times 2.64\,k}{33 - 3.79} = 0.324\,k\Omega$$

5. Following the recommendations published in the tube data sheet, the operating conditions of the lower triode can be chosen, for example : **$U_{a1} \approx 80$ V**, **$I_a \approx 9.5$ mA** and **$U_g = -1.5$ V**

   $U_g = -1.5V$, → **$R_{k1} = Ug / I_a = 1500\ mV / 9.5\ mA = 0.1579\ k\Omega$**. (Standard: **$R_{k1} = 160$** Ω)



6. For the upper tube:
   Draw the upper tube (V_2) load line with a slope of $1 / R_{a1}$ (1 / 9.1 kΩ). In general, the load line can be placed anywhere on the $I_a = f(U_a)$ diagram (in the allowable area of tube operation). But, since the equivalent current flows through both tubes, the quiescent point must be located at the intersection of the load line and the line I_a = current flowing through both tubes, i.e. I_a = 9.5 mA. If the load line intersects the line I_a = 9.5 mA at a point different from the quiescent point of tube V_1, the operating conditions of the upper tube (V_2), U_a and U_g, will differ from the operating conditions of tube V_1.
   The load line of the tube V_2 intersects the abscissa axis at the point $\mathbf{U_{ab(V2)}}$ **,** so the total voltage $\mathbf{U_{ab}}$ is:
   $\mathbf{U_{ab} = U_{a(V1)} + U_{ab(V2)}}$

Requirements for specific operating conditions of the tube or the availability of the voltage power supply can be fulfilled by moving the load line of the tube V_2 within the allowable operating area of the tube on the diagram $I_a = f(U_a)$, but so that the quiescent point must be located at the intersection of the load line with the I_a line – current flowing through both tubes. In this example, the choice is that both tubes have equal quiescent points, so the quiescent point is: **M (80 V, 9.5 mA),** $\mathbf{U_g = -1.5\ V}$. The load line of the tube V_2 intersects the abscissa at the point $\mathbf{U_{ab(V2)} = 170\ V}$, so the total power supply voltage is:

$$\mathbf{U_{ab}} = U_{a(V1)} + U_{ab(V2)} = 80\text{ V} + 170\text{ V} = \mathbf{250\ V}$$

7. Calculation of R_1 and R_2

$\mathbf{U_1 \approx U_{a1} = 80\ V}$

$$U1 \approx U_{a1} = 80V = \frac{R_2}{R_1+R_2} \times U_{ab} \rightarrow \frac{R_2}{R_1} = \frac{U_1}{U_{ab}-U_1} = \frac{80V}{250V-80V} = 0.47$$

The combination of R_1 and R_2 that fulfills the condition $(R_2 / R_1) = 0.47$: $\mathbf{R_1 = 464\ k\Omega}$, $\mathbf{R_2 = 220\ k\Omega}$

$$(U1) = \frac{R_2}{R_1+R_2} \times U_{ab} = \frac{220k}{464k+220k} \times 250V = 80.4V$$

a) The grid of the upper tube (V2) must be grounded for the AC signal. Therefore, C_2 must be connected across R_1: C_2 = 0.1μF.

b) Calculation of capacitor C_k connected across the V1 cathode resistor R_{k1}:
   The resistance of the grounded cathode circuit (excluding the cathode resistor) seen from the cathode is:

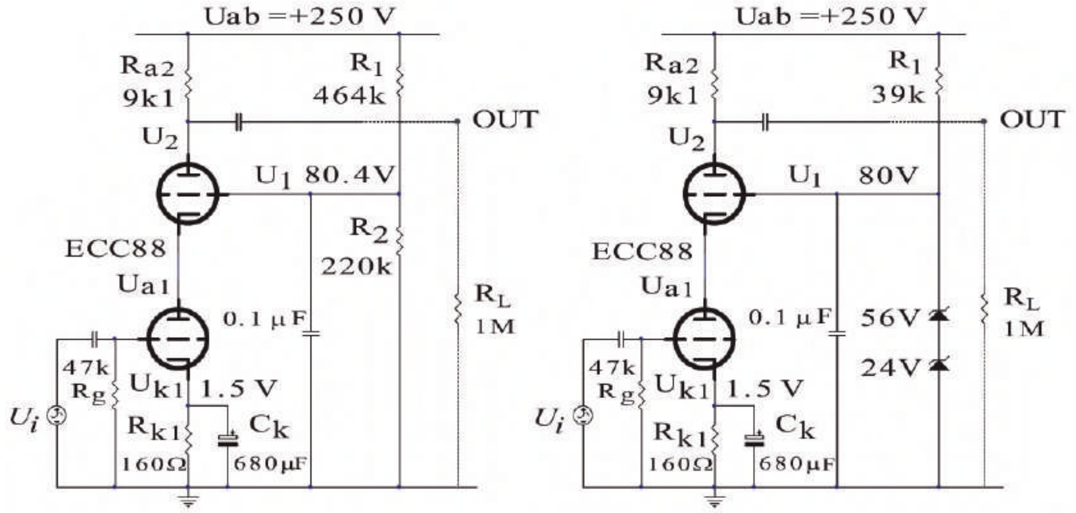
$$R_k' = \frac{R_{i1}+R_{a1}}{\mu+1} = \frac{2.64k+0.342k}{33+1} = 0.0877k\Omega \qquad R_{keq} = R_k' \parallel R_k = \frac{0.0877k \times 0.16k}{0.0877k+0.16k} = 0.057k\Omega$$

If the required cutoff frequency is: $\mathbf{f_{low(-3dB)}}$ **=** 5Hz:

$$C_k = \frac{1}{2\times\pi\times R_{keq}\times f_{L(-3dB)}} = \frac{1}{2\times 3.14\times 57\Omega\times 5Hz} = 558.7\mu F$$

Standard: 680 μF

Complete design:



 # Any source of stable DC voltage can be used to create the grid voltage U_1 of the upper tube.

For example, it can be designed using a Zener diode: $V_Z = U_1$ and $I_Z = (U_{ab} - V_Z) / R_1$
In this example:
$V_Z = U_1 = 80$ V
$I_Z = (U_{ab} - V_Z) / R_1 = (250\text{ V} - 80\text{ V}) / 39\text{ k} = 4.36$ mA

4.5 SRPP (Shunt Regulated Push–Pull)

BASIC CIRCUIT

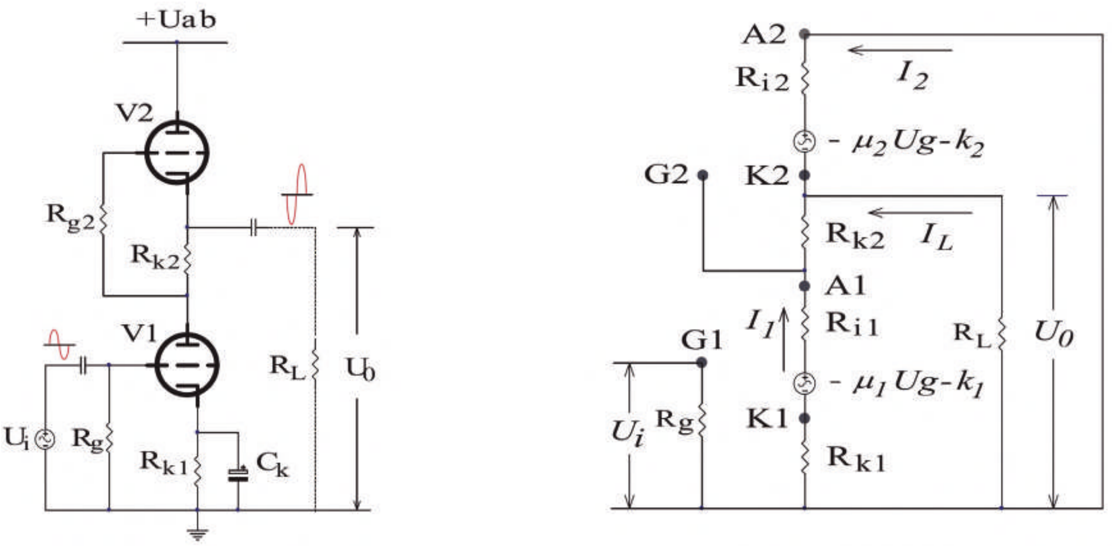


Fig. 4-15

The lower triode operates as a grounded cathode amplifier with an active load. The current of the signal flowing through the lower triode also flows through the cathode resistor Rk2 of the upper triode and causes a change of voltage across R_{k2}, i.e. the signal appears across R_{k2}. As R_{k2} is connected between the grid and the cathode of the upper triode, the signal voltage also appears between the grid and the cathode of the upper triode. When a negative signal is applied to the input of the lower triode, the lower triode becomes less conductive, the current through the lower triode and through the cathode resistor R_{k2} of the upper triode decreases and the voltage drop across R_{k2} also decreases. As the voltage drop across R_{k2} decreases, the upper triode becomes more conductive and conducts more current. In a similar way, the situation can be described when a positive signal is applied to the grid of the lower triode, but the result is the opposite of that obtained when a negative signal is applied to the grid of the lower triode. Thus, the lower and upper triodes operate in opposite phases, similar to the operation of a push-pull amplifier. The inequality of the currents flowing through the lower and upper triodes caused by the unequal conductivity of the lower and upper triodes is compensated by the charging and discharging currents of the coupling capacitor.

A few important notes about the SRPP amplifier:

- **The SRPP amplifier can only operate in class A.**
- **The SRPP amplifier only operates properly if it is loaded.**
- **The output signal of the SRPP amplifier is generated at the cathode of the upper triode.**
- **The lower and upper triodes operate in opposite phases -** similar to the operation of a push – pull amplifier. This is the reason for the name of this amplifier circuit: *Shunt Regulated Push - Pull* or *SRPP*.

Analyses of SRPP amplifier using Fig.4-15: the grid – cathode voltage is equal to the input voltage plus the voltage across the cathode resistor $U_{gk1} = U_i + I_1 \times R_{k1}$ and $U_{gk2} = I_1 \times R_{k2}$.

The output voltage seen from the lower triode considering R_{k2} is:

$$U_0 = -\mu_1 \times U_{gk1} - (R_{k1} + R_{i1} + R_{k2}) \times I_1 = -\mu_1 \times U_i - (R_{k1} + R_{k2} + R_{i1}) I_1$$

$$U_0 = -\mu_1 \times U_i - (R_{k1} \times (\mu_1 + 1) + R_{i1} + R_{k2}) \times I_1 \tag{401.39}$$

The output voltage seen from the upper triode is:

$$U_0 = \mu_2 \times U_{gk2} - R_{i2} \times I_2 = \mu_2 \times R_{k2} \times I_1 - R_{i2} \times I_2 \tag{401.40}$$

Also:

$$U_0 = -R_L \times I_L \tag{401.41}$$

and

$$I_1 + I_2 + I_L = 0 \tag{401.42}$$

Solving the above system of four equations, the **amplification (**A = U_o / U_i) is:

CASE 1. The general amplification equation:

$$A = -\mu_1 \times \frac{\mu_2 \times R_{k2} + R_{i2}}{(\mu_1 + 1) \times R_{k1} + R_{i1} + (\mu_2 + 1) \times R_{k2} + R_{i2} + \frac{R_{i2} \times [(\mu_1 + 1) \times R_{k1} + R_{i1} + R_{k2}]}{R_L}}$$

CASE 1a. $R_L \longrightarrow \infty$ (unloaded amplifier, amplifier without R_L):

$$A = -\mu_1 \times \frac{\mu_2 \times R_{k2} + R_{i2}}{(\mu_1 + 1) \times R_{k1} + R_{i1} + (\mu_2 + 1) \times R_{k2} + R_{i2}}$$

CASE 1b. Lower triode cathode resistor R_{k1} bypassed by capacitor C_{k1} :

$$A = -\mu_1 \times \frac{\mu_2 \times R_{k2} + R_{i2}}{R_{i1} + (\mu_2 + 1) \times R_{k2} + R_{i2} + \frac{R_{i2} \times (R_{i1} + R_{k2})}{R_L}}$$

CASE 1b1. $R_L \longrightarrow \infty$ (unloaded amplifier, amplifier without R_L):

$$A = -\mu_1 \times \frac{\mu_2 \times R_{k2} + R_{i2}}{R_{i1} + (\mu_2 + 1) \times R_{k2} + R_{i2}}$$

In practical application, the lower and the upper triode of the same type are used and both triodes have an identical bias voltage, i.e. $\mu_1 = \mu_2 = \mu$, $R_{i1} = R_{i2} = R_i$, $R_{k1} = R_{k1} = R_k$.

CASE 2.

$$A=-\mu\times\frac{\mu\times R_k+R_i}{2\times(\mu+1)\times R_k+2\times R_i+\dfrac{R_i\times[(\mu+2)\times R_k+R_i]}{R_L}}$$

CASE 2a. $R_L \longrightarrow \infty$ (unloaded amplifier, amplifier without R_L):

$$A=-\mu\times\frac{\mu\times R_k+R_i}{2\times(\mu+1)\times R_k+2\times R_i}$$

CASE 2b. Lower triode cathode resistor bypassed by capacitor C_{k1} :

$$A=-\mu\times\frac{\mu\times R_k+R_i}{2\times R_i+(\mu+1)\times R_k+\dfrac{R_i\times(R_i+R_k)}{R_L}}$$

CASE 2b1. $R_L \longrightarrow \infty$ (unloaded amplifier, amplifier without R_L):

$$A=-\mu\times\frac{\mu\times R_k+R_i}{2\times R_i+(\mu+1)\times R_k}$$

Output impedance Z_{out}:

CASE 1. General equation:

$$Z_{out}=\frac{R_{i2}\times[(\mu_1+1)\times R_{k1}+R_{i1}+R_{k2}]}{(\mu_1+1)\times R_{k1}+R_{i1}+(\mu_2+1)\times R_{k2}+R_{i2}}$$

CASE 2. The lower triode cathode resistor bypassed by capacitor C_{k1} :

$$Z_{out}=\frac{R_{i2}\times(R_{i1}+R_{k2})}{R_{i1}+(\mu_2+1)\times R_{k2}+R_{i2}}$$

CASE 3. Both triodes are identical and have equal grid biases:

$$Z_{out}=\frac{R_i\times[(\mu+2)\times R_k+R_i]}{2\times(\mu+1)\times R_k+2\times R_i}$$

CASE 3a. The lower triode cathode resistor bypassed by capacitor C_{k1} :

$$Z_{out}=\frac{R_i\times(R_i+R_k)}{2\times R_i+(\mu+1)\times R_k}$$

CASE 4. Upper and lower triode: the same triode type, equal grid biases of upper and lower triode , equal cathode resistors of upper and lower triode (R_k) , R_k of lower tube bypassed by capacitor:

$$A=\frac{-\mu}{2} \quad \text{and} \quad R_{out}=\frac{R_i}{2}$$

 # Note:

Analyzing the above equations, it can be concluded that the effects of ***bypassing the cathode resistor of the lower triode*** *are:*

Higher amplification
Lower output impedance.

	R_K		R_L						Z_{out}
			∞	1M	470k	100k	50k	10k	
ECC83	1k	A	49.69	48	46.5	37.8	30.5	-	31.44k
		A_{Ck1}	71.9	70.6	69.3	61.16	53.2	-	17.56k
	2k2	A	49.61	48	46.49	37.7	30.4	-	31.49k
		A_{Ck1}	81.36	80.4	79.39	72.87	65.99	-	11.64k
ECC82	1k2	A	8.15	8.11	8.08	7.83	7.54	5.82	4k
		A_{Ck1}	12.9	12.87	12.84	12.66	12.43	10.88	1.85k
ECC81	1k	A	29.58	29.4	29.2	28	26.6	18.99	5.57k
		A_{Ck1}	51.32	51.23	51.14	50.5	49.65	44.2	1.59k
ECC88	200Ω	A	16.15						1.35k
		A_{Ck1}	25.23						0.624k
6SN7	1k	A	9.65						3.98k
		A_{Ck1}	15.22						1.688k
6SL7	1k	A	34.7						22.19k
		A_{Ck1}	50.19						12.45k
12BH7	2k	A	8.09						2.882k
		A_{Ck1}	14.29						0.877k

Conclusion

Main characteristics of SRPP amplifier:

- Input and output signals are **opposite** in phase
- low output impedance
- low harmonic distortion
- good amplitude characteristic

#Note: The lower and upper triodes do not have to be of the same type. Also, it is not necessary that the lower and upper triodes have equal grid biases (but the quiescent currents must be equal).

Optimized SRPP

"Making both tubes contribute equally to the load current, and also maximize that current" Merlin Blencowe, *Audio Xpress*, June 2010. Mr. Blencowe's analysis:

Lower tube cathode resistor bypassed by C_{k1}.

Quiescent current through the triode is (Equation 307): $I_0=\frac{U_{ab}}{\mu\times R_k+R_i}$

In SRPP configuration two tubes are connected in series, so: $I_0=\frac{U_{ab}}{2\times\mu\times R_{k2}+2\times R_i}$ *(Eq.I)*

*The circuit is **balanced** when the quiescent current is **half** the peak current through the tube. If the U_{ab} is shared equally between two triodes then the max. current through the lower triode is:* $\frac{\frac{U_{ab}}{2}}{R_i+R_L+R_{k2}}$ *while for the upper triode it is:* $\frac{\frac{U_{ab}}{2}}{R_i+R_L}$*. Since they are not quite balanced, it is necessary to add an additional resistor **R_a equal to R_{k2}** in series* *with upper triode. If* $R_{k1} = R_{k2} = R_a = R_k$ *, the peak current for either tubes is:* $I_{peak}=\frac{U_{ab}}{2\times(R_i+R_L+R_k)}$. *For proper balance, the quiescent current must be half of the peak current:* $I_0=\frac{U_{ab}}{4\times(R_i+R_L+R_k)}$ *(Eq. II)*

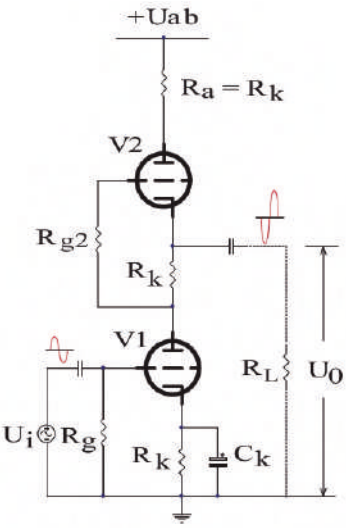
Quiescent currents are equal if Eq. I (with added $R_a= R_k$*) = Eq. II:* $4\times(R_i+R_L+R_k)=2\times\mu\times R_{k2}+2\times R_i$

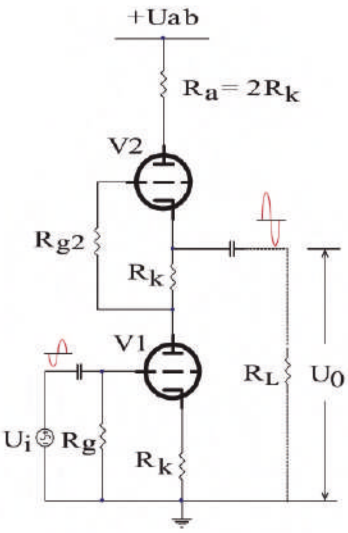
Solving for R_k :

$$R_{k\,optimum}=\frac{2\times R_L+R_i}{\mu-1.5} \quad or \quad R_L=\frac{R_k\times(\mu-1.5)-R_i}{2}$$

The amplification is:

$$A=-\mu_1\times\frac{R_L\times(R_i+R_a+\mu_2\times R_k)}{(R_{i1}+R_k)\times(R_{i2}+R_a)+R_L\times[R_{i1}+R_{i2}+R_a+R_k\times(\mu_2+1)]}$$





If $R_{k1} = R_{k2} = R_a = R_k$:

$$A=-\mu\times\frac{R_L\times(R_i+(\mu+1)\times R_k)}{(R_i+R_k)^2+R_L\times[2\times R_i+R_k\times(\mu+1)]}$$

The output impedance Z_{out} is:

$$Z_{out}=\frac{(R_{i2}+R_a)\times(R_{i1}+R_k)}{R_{i1}+R_{i2}+R_a+(\mu_2+1)\times R_k}$$

or simplified:

$$Z_{out}=\frac{(R_i+R_k)^2}{2\times R_i+(\mu+2)\times R_k}$$

Conclusion:
There is always an optimal R_k ***for any*** R_{Load} ***.***

If R_{k1} ***is <u>not bypassed</u>,*** $R_a = 2R_k$

$$R_{k\,optimum}=\frac{2\times R_L+R_i}{\mu-1} \quad \text{and} \quad I_0=\frac{U_{ab}}{2\times[R_i+(\mu+2)\times R_k]}$$

Amplification is:

$$A=-\mu_1\times\frac{R_L\times(R_{i2}+R_a+\mu_2\times R_{k2})}{[R_{i1}+R_{k2}+R_{k1}\times(\mu_1+1)]\times(R_{i2}+R_a)+R_L\times[R_{i1}+R_{i2}+R_a+R_{k1}(\mu_1+1)+R_{k2}\times(\mu_2+1)]}$$

Output impedance Z_{out} ***(***$R_{k1} = R_{k2} = R_a = R_k$ *,* $\mu_1 = \mu_2 = \mu$*,* $R_{i1} = R_{i2} = R_i$***) is:***

$$Z_{out}=\frac{(R_i+2\times R_k)\times[R_i+R_k\times(\mu+2)]}{2\times[R_i+R_k\times(\mu+2)]}$$

4.6 THE μ FOLLOWER

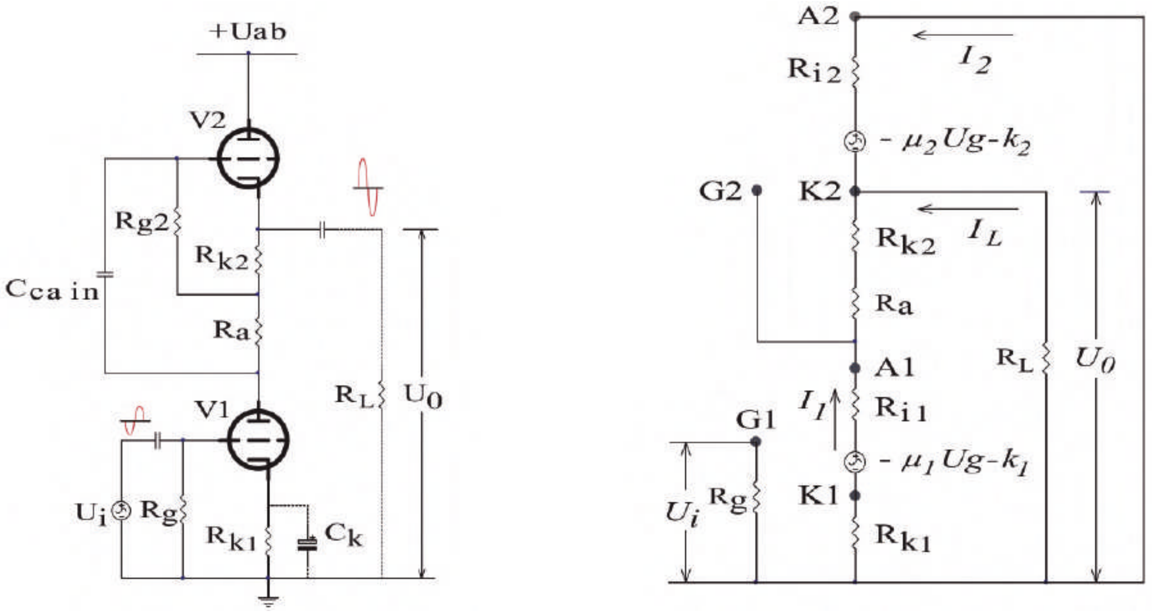


Fig. 4-16

The μ follower is a combination of two basic topologies of tube amplifiers:

- The lower triode circuit is a grounded cathode amplifier
- The upper triode circuit is a grounded anode amplifier and acts as an active load of the lower tube (together with load resistor).

The signal generated at the anode of the lower tube is fed (via the capacitor $C_{ca\ in}$) to the grid – the input of the upper tube (grounded anode amplifier). The gain of the grounded anode amplifier is nearly equal to unity gain so the output signal at the cathode of the grounded anode amplifier (upper tube) is nearly equal to that at the anode of the lower tube. Therefore, there is no signal loss on the anode load of the lower tube, i.e. the circuit of the upper tube (grounded anode amplifier) acts as a constant current source.

The input impedance of the grounded anode amplifier is very high $R_{in} = (\mu + 1) \text{ x } (R_{g2} + R_{k2})$ and the amplification of the lower tube is close to the μ of the lower tube (name: μ follower).

Taking into account: $U_{gk1} = U_i + I_1 \times R_{k1}$, $U_{gk2} = I_1 \times R_{k2}$ and $R_{k2\ tot} = R_{k2} + R_a$, the **amplification** is:

$$A = A_{grounded\,cathode} \times A_{grounded\,anode} \approx \mu \times \frac{\mu}{(1+\mu) + \frac{R_i}{R_{k2}+R_a}} \qquad (401.43)$$

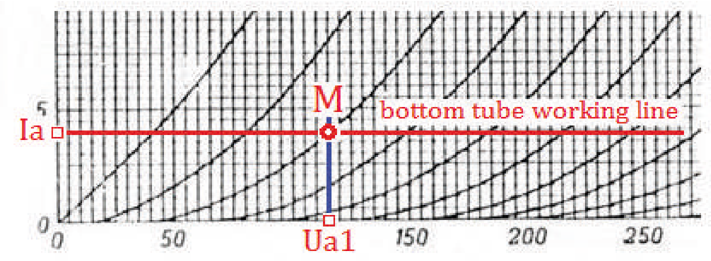
and **output impedance:**

$$R_{out} = \frac{R_i \times [R_i + (\mu+1) \times R_{k1} + R_{k2} + R_a]}{2 \times R_i + (\mu+1) \times (R_{k1} + R_k + R_a)} \qquad (401.44)$$

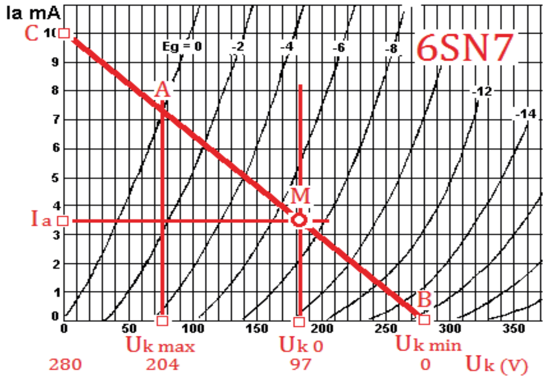
If **$R_{k1} = R_{k2} = R$**:

$$R_{out} = \frac{R_i \times [R_i + (\mu+2) \times R_k + R_a]}{2 \times [R_i + (\mu+1) \times R_k] + (\mu+1) \times R_a} \qquad (401.45)$$

Graphical analysis:
The lower triode is loaded with a constant current source, so its load line is a straight line parallel to the U_a axis of the anode characteristics diagram. The chosen I_a, i.e. I_a = constant and U_a (U_g), determine the quiescent point located on the load line.



The upper triode circuit is a grounded anode amplifier and graphical analysis of the grounded anode amplifier can be applied, but with the **condition** that **the current of the upper triode must be equal to the current of the lower triode** at the quiescent point.



So, it is necessary to draw the load line of the grounded anode amplifier with a slope of 1 / (R_{k2}+R_a) and a supply voltage equal to the total supply voltage minus U_a of the lower triode. The quiescent point is located at a point located on the load line with I_a equal to I_a of the lower triode.

#Note

The lower and upper triodes do not have to be of the same type. Also, it is not necessary that the lower and upper triodes have identical quiescent points (U_a and U_g can be different, but the quiescent currents I_a of both tubes must be equal).

Main characteristics of μ follower:

- Input and output signals are **opposite** in phase.
- low output impedance
- low harmonic distortion
- good amplitude characteristic

Example 406.1:

μ follower amplifier circuit.
Tube: ECC82 (R_i = 7.7 kΩ, μ = 18)
Available power supply: U_{ab} = 310 V

1. *The quiescent point of the lower triode is chosen to be: U_a = 105 V, I_a = 3.5 mA and U_g = -3.5 V. The load line of the lower triode is a straight line parallel to the U_a axis i.e. I_a = Constant = 3.5 mA. It intersects the characteristic U_g = -3.5V at point M. The coresponding U_a is: U_a = 105 V.*
2. *The cathode resistor of the lower triode is: $R_{k1} = U_g / I_a$ = 3.5 V / 3.5 mA = 1 kΩ.*
3. *The upper triode circuit is the cathode follower amplifier. The "power supply" of the upper triode is the total power supply voltage minus U_a (≈ U_{ak}) of the lower triode: $U = U_{ab} - U_{ak\ \text{(lower triode0}}$ = 310 V – 105 V = 205 V. In order to fulfill the condition that equal currents flow through the lower and upper triode, it is necessary to set the load line of the upper triode by drawing a straight line through the points X (205 V) on the U_a axis and M – quiescent point (if it is chosen that the quiescent points of the upper and lower triodes are equal) up to the intersection with the I_a axis, point Y(7.3 mA).*

The total resistance in the cathode circuit of the cathode follower ($R_{k2} + R_a$) is determined by the slop of the load line: ($R_{k2} + R_a$) = $U_{\text{at the point X}}$ / $I_{a\ \text{at the point Y}}$ = 205 V / 7.3 mA = 28 kΩ.

The quiescent point of the upper triode is:
U_a = 105 V, I_a = 3.5 mA and U_g = -3.5 V, so the $R_{k2} = U_g / I_a$ = 3.5 V / 3.5 mA = 1 kΩ.
R_a = 28 kΩ – R_{k2} = 28 kΩ – 1 kΩ = 27 kΩ.

Note 1#:

In the example above, the same type of triode and equal quiescent point are chosen for the upper and lower triodes, but this is not necessary. The quiescent points of the upper and lower triodes may be different, but the quiescent currents of both tubes must be equal. In that case, the quiescent point of the upper triode must be set at the intersection of its load line and the line I_a = constant = the quiescent current of the lower triode parallel to U_a axis - (the intersection of the load line and line $I_a = I_{a\ \text{quiescent (lower triode)}}$ = constant determines the quiescent point of the upper triode ($U_{a\ \text{upper triode}}$, $U_{g\ \text{upper triode}}$). The load line of the upper triode intersects the I_a axis at *point Y). The slope of the load line of the upper tube is determined by the total cathode resistance (R_{k2} + R_a): 1 / (R_{k2} + R_a)*

Note 2#:

In the example above, to make the design easier, knowledge and experience from previous analyzes of basic amplifier circuits is used: it is chosen that U_a at a quiescent point of the lower triode is equal to 1/3 of U_{ab} (approximately). Thus, to obtain the maximum cathode follower amplifier headroom (maximum output voltage), bearing in mind that the "supply voltage" of the upper triode is equal to 2/3 U_{ab} (approximately), the U_a of the upper tube is chosen to be equal to 1/2 "supply voltage" of the upper tube (approximately), i. e. $U_{a\ upper\ triode} \approx \frac{1}{2}\left(U_{ab} - U_{a\ lower\ triode}\right)$.

1. *Amplification*

$$A = A_{grounded\,cathode} \times A_{grounded\,anode} \approx \mu \times \frac{\mu}{(1+\mu)+\frac{R_i}{R_{k2}+R_a}} = 18 \times \frac{18}{(1+18)+\frac{7.7}{1+27}} = 16.8$$

2. *Output impedance*

$$R_{out} = \frac{R_i \times \left[R_i + (\mu+2) \times R_k + R_a\right]}{2 \times \left[R_i + (\mu+1) \times R_k\right] + (\mu+1) \times R_a} = \frac{7.7 \times \left[7.7 + (18+2) \times 1 + 27\right]}{2 \times \left[7.7 + (18+1) \times 1\right] + (18+1) \times 27} = 0.743\,k\Omega$$

3. *Upper triode (Cathode follower)*

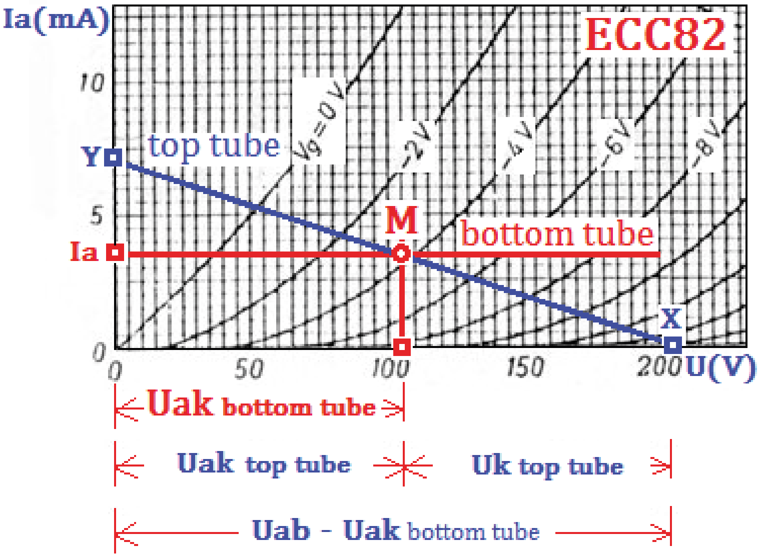
- *Amplification:* $A_{CF} = \frac{\mu}{(\mu+1)+\frac{R_i}{R_{k2}+R_a}} = \frac{18}{(18+1)+\frac{7.7}{1+27}} = 0.933$

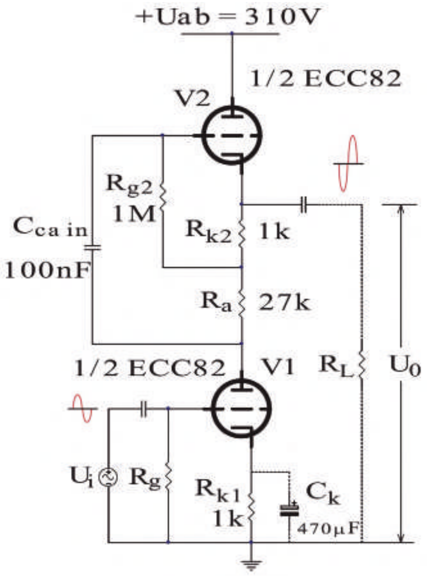
- *Input impedance:*

$$R_{in\ CF} = (\mu + 1) \times \left(R_{g2} + R_a\right) = (18 + 1) \times (1000 + 27) = 19.5\ M\Omega$$

4. *AC load of the lower triode:*

$$R_{L\,lower\,triode} = \frac{(R_{k2}+R_a)}{1 - A_{CF(upper\,triode)}} = \frac{(1+27)}{1-0.933} = 417.9\,k\Omega$$





4.7 DIFFERENTIAL AMPLIFIER

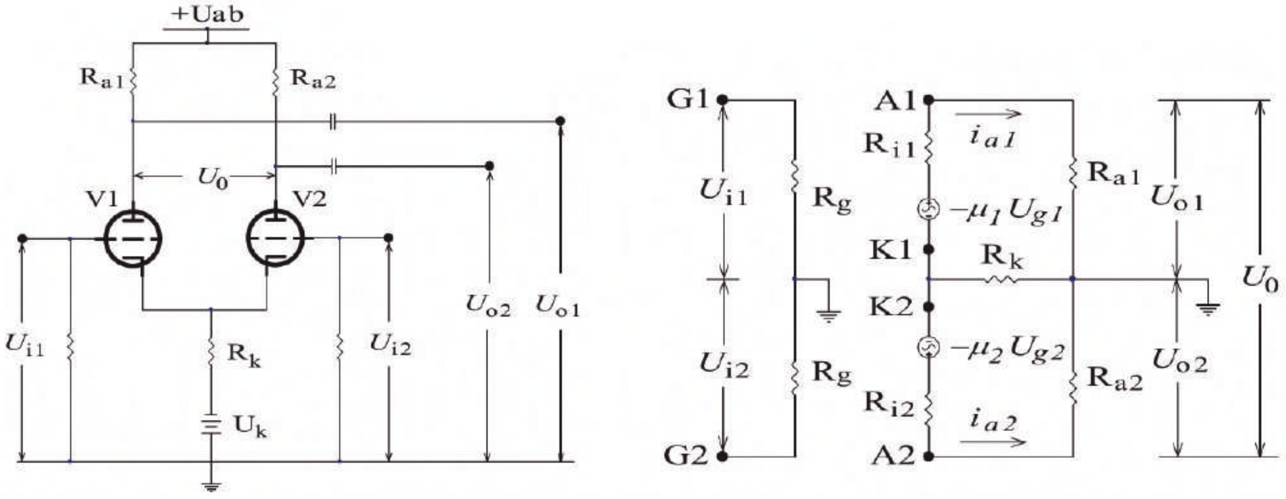


Fig. 4-17

Using the equivalent circuit:

$$-\mu \times u_{g1} = R_{i1} \times i_{a1} + R_{a1} \times i_{a1} + R_k \times (i_{a1} + i_{a2})$$
$$-\mu \times u_{g2} = R_{i2} \times i_{a2} + R_{a2} \times i_{a2} + R_k \times (i_{a1} + i_{a2})$$

Also:

$$u_{g1} = u_{i1} + R_k \times (i_{a1} + i_{a2)})$$
$$u_{g2} = u_{i2} + R_k \times (i_{a1} + i_{a2)})$$

If the triodes are the same type, i.e., $\mu_1=\mu_2=\mu$, $R_{i1}=R_{i2}=R_i$, $R_{a1}=R_{a2}=R_a$, and $u_{a1} = R_a \times i_{a1}$, $u_{a2} = R_a \times i_{a2}$, by solving previous equations, i_{a1} and i_{a2}:

$$i_{a1}=\mu \times \frac{u_{i2}-\left[1+\frac{(R_i+R_a)}{R_k \times (\mu+1)}\right] \times u_{i1}}{\left\{\left[1+\frac{(R_i+R_a)}{R_k \times (\mu+1)}\right]^2-1\right\} \times R_k \times (\mu+1)}$$

$$i_{a2}=\mu \times \frac{u_{i1}-\left[1+\frac{(R_i+R_a)}{R_k \times (\mu+1)}\right] \times u_{i2}}{\left\{\left[1+\frac{(R_i+R_a)}{R_k \times (\mu+1)}\right]^2-1\right\} \times R_k \times (\mu+1)}$$

The output voltage, anode (V1) - anode (V2) (voltage between the anode of tube V1 and the anode of tube V2), is:

$$U_0=R_a \times i_{a1}-R_a \times i_{a2}=R_a \times (i_{a1}-i_{a2})$$

By substituting i_{a1} and i_{a2} in the above equation:

$$U_0=\mu \times \frac{2+\frac{R_i+R_a}{R_k \times (\mu+1)}}{\left(1+\frac{R_i}{R_a}\right) \times \left(2+\frac{R_i+R_a}{R_k \times (\mu+1)}\right)} \times (u_{i1}-u_{i2}) \qquad (401.46)$$

By analyzing the Equation (401.46), it can be concluded that the differential amplifier:

1. **amplifies only the difference between the input signals ($u_{i1} - u_{i2}$)**
2. **rejects any voltage signals that are common to both inputs ($u_{i1} = u_{i2}$)**

   (Common Mode Rejection Ratio $CMRR \approx \frac{\mu \times R_k}{R_i+R_a}$).

Power Supply Rejection Ratio PSRR: *Hum and noise generated by the power supply are present at the inputs of both triodes (common signals) and are rejected by the differential amplifier. Also, the differential amplifier attenuates the power supply hum:* $A_t = \frac{R_a + 2 \times R_i}{2 \times (R_a + R_i)}$*. So, the total rejection ratio of unwanted signals originating from the power supply (Power Supply Rejection Ratio) is the sum of CMRR and attenuation* A_t *:*

$$PSRR \approx \frac{\mu \times R_k}{R_i + R_a} + \frac{R_a + 2 \times R_i}{2 \times (R_a + R_i)} = \frac{R_a + 2 \times (R_i + \mu \times R_k)}{2 \times (R_i + R_a)}$$

- **Amplification of the signal applied to the triode grid:**

$$A_{11} = \frac{-\mu \times R_a}{(R_i + R_a) \times [2 + \frac{R_i + R_a}{R_k \times (\mu + 1)}]} \times [1 + \frac{R_i + R_a}{R_k \times (\mu + 1)}]$$

- **Amplification of the triode of the signal applied to the grid of the other triode:**

$$A_{12} = \frac{\mu \times R_a}{(R_i + R_a) \times [2 + \frac{R_i + R_a}{R_k \times (\mu + 1)}]}$$

- **The input and output signals of each triode are opposite in phase.**

- **The input signal of one triode and the output signal of the other triode are in phase.**

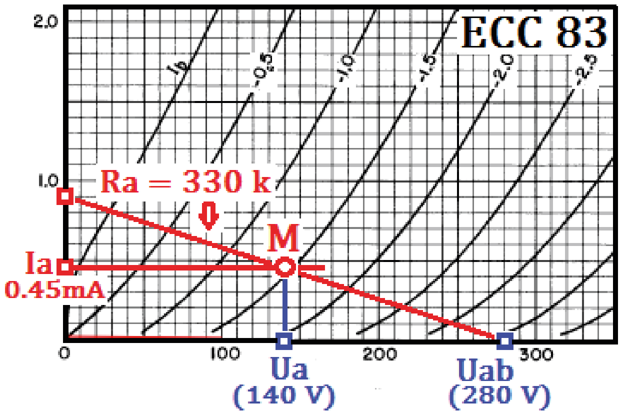
- **The amplification of one triode and the amplification of the other triode are not equal (this difference is:** $\frac{A_{11}}{A_{12}} = 1 + \frac{R_i + R_a}{R_k \times (\mu + 1)}$ **and it is less if (μ + 1) x** R_k » (R_i **+** R_a**)***).

  (In order to equalize the amplification of both triodes, it is necessary to reduce the resistance of the anode resistor of the first tube:)

$$R_{a1} = \frac{R_a}{1 + \frac{R_i + R_a}{R_k \times (\mu + 1)}}$$

* *In order to fulfill the condition* **(μ+1) x** R_k » **(**R_i**+**R_a**),** *in practice constant- current sinks can be used in cathode circuits (instead* R_k *), as well as triodes with high μ:*

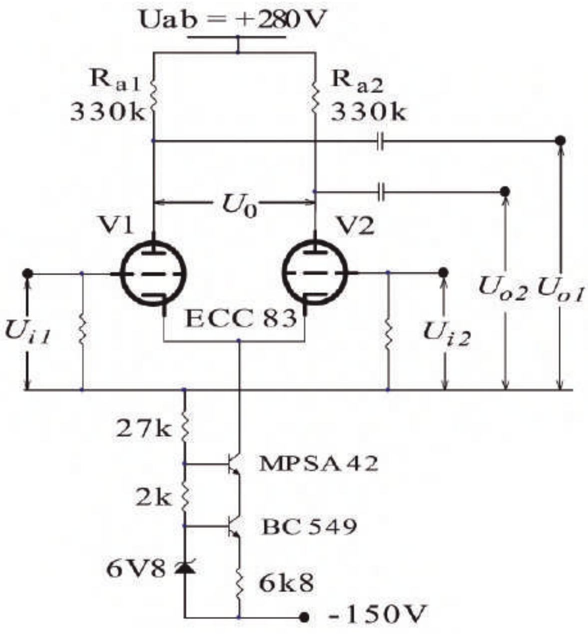
Example:
A high μ triode: ECC 83 (μ = 100)
Power supply = +280V, R_a *= 330 kΩ, quiescent point M:* U_a *= 140V,* I_a *= 0.45mA, and* U_g *= - 1.4V.*
So, it is necessary to design a constant current sink: I = 2 x I_a *= 2 x 0.45 mA = 0.9 mA*



The current of the constant current sink circuit (Zener diode 6V8, R_e *= 6k8, transistor BC 549 (*h_{fe} *= 200)), is:*
I = (U_Z *–* U_{b-e} *) /* R_e *= (6.8 – 0.7) / 6k8 = 0.897mA.*

Considering that the auxiliary power supply – 150V is used, the current flowing through the Zener diode is limited by resistors 27k and 2k to the I_Z *= 150V / 29k = 5.17mA.*
The high voltage transistor is MPSA 42 (h_{fe} *=40).*
The resistance of the constant current sink is:

$$R_{out\,CCS} \approx R_e \times h_{fe(T1)} \times h_{fe(T2)} = 6k8 \times 200 \times 40 = 54.4\ M\Omega$$



Since the resistance of the constant current sink $R_{\text{out CCS}}$ is very high, the expression $\frac{R_i+R_a}{R_k\times(\mu+1)}=\frac{R_i+R_a}{R_{outCCS}\times(\mu+1)}$

in the equations A_{11} and A_{12} is very small and negligible, equations A_{11} and A_{12} (amplifications) are equal:

$$A_{11}=A_{12}=\frac{\mu\times R_a}{(R_i+R_a)\times 2}=\frac{100\times 330k}{2\times(62.5k+330k)}=42(32.47\,dB)$$

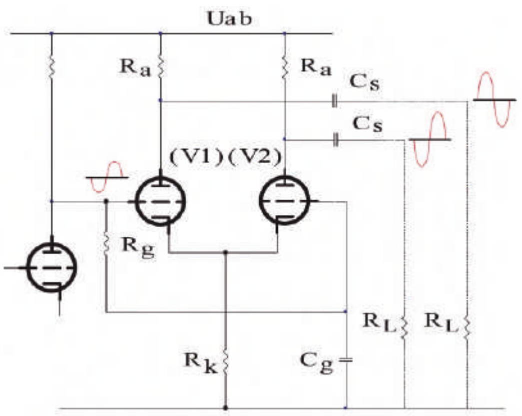
The voltage amplification, anode (V1) – anode (V2) (voltage between the anode of tube V1 and the anode of tube V2), is:

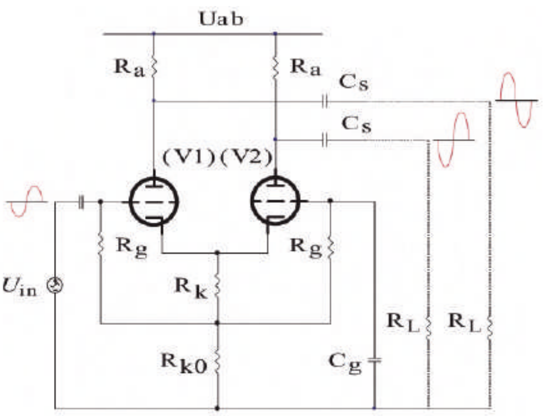
$$\frac{u_0}{(u_{i1}-u_{i2})}=\mu\times\frac{2+\frac{R_i+R_a}{R_k\times(\mu+1)}}{\left(1+\frac{R_i}{R_a}\right)\times\left(2+\frac{R_i+R_a}{R_k\times(\mu+1)}\right)}=\mu\times\frac{R_a}{R_i+R_a}=100\times\frac{330k}{62.5k+330k}=84$$

PSRR:

$$PSRR\approx\frac{R_a+2\times(R_i+\mu\times R_k)}{2\times(R_i+R_a)}=\frac{330k+2\times(62.5k+100\times 54400k)}{2\times(62.5k+330k)}=13860=82.8\,dB$$

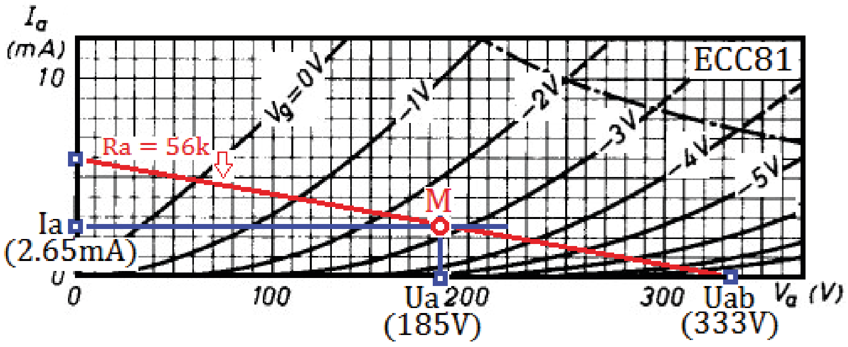
The differential amplifier circuit is easily reconfigured and used as a phase splitter circuit





Example 407.1:

Phase splitter amplifier circuit.
Tube: ECC81: R_i = 16; μ = 60
The cathodes have to be elevated to a potential of 67 V (approximately).
Available power supply: U_{ab} = 400 V

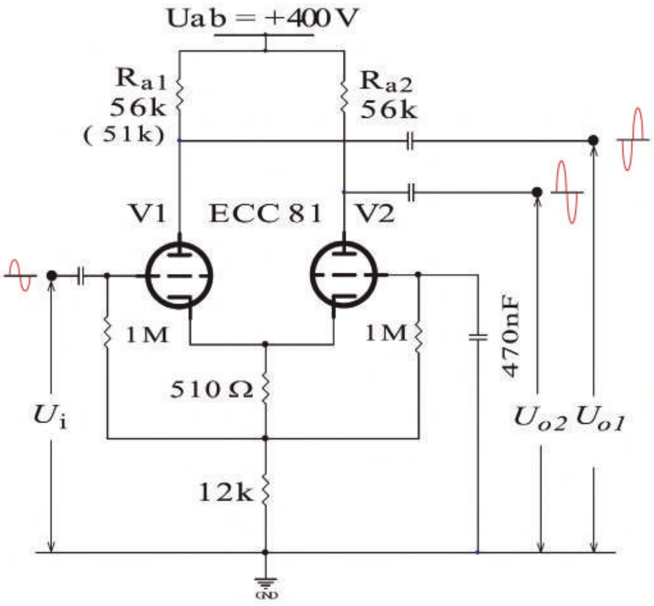


1. Choose the quiescent point somewhere in the middle of the anode characteristic diagram, for example: M (U_a = 185 V, I_a = 2.65 mA) and U_{gk} = - 2.7 V. To design the load line, use: $U_{ab'} = U_{ab} - U_k$ = 400 V – 67 V ≈ 333 V.
   Draw the load line through the points $U_{ab'}$ = 333 V and the quiescent point M. It intersects the I_a axis at the point I_a = 5.9 mA. The anode load is $R_a = U_{ab'}$ / 5.9 mA = 333 V / 5.9 mA = 56.44 kΩ ≈ 56 kΩ
2. $R_k = U_{gk}$ / 2 x I_a = 2.7 V / 2 x 2.65 mA = 0.509 kΩ ≈ 0.51 kΩ
3. R_{k0} = (U_k - U_{gk}) / 2 x I_a = (67 V – 2.7 V) / 2 x 2.65 mA = 12.13 kΩ ≈ 12 kΩ
4. Amplification of inverted A_{11} and amplification of the non-inverted output A_{12}:

$$A_{11} = -\frac{\mu \times R_a}{(R_i + R_a) \times \left[2 + \frac{R_i + R_a}{R_{kt} \times (\mu + 1)}\right]} \times \left[1 + \frac{R_i + R_a}{R_{kt} \times (\mu + 1)}\right]$$

$$= \frac{60 \times 56k}{(16k + 56k) \times [2 + \frac{16k + 56k}{12.51k \times (60 + 1)}]} \times \left[1 + \frac{16k + 56k}{12.51k \times (60 + 1)}\right] = 24.38 = 27.74\ dB$$

$$A_{12} = \frac{\mu \times R_a}{(R_i + R_a) \times [2 + \frac{R_i + R_a}{R_{kt} \times (\mu + 1)}]} = \frac{60 \times 56k}{(16k + 56k) \times [2 + \frac{16k + 56k}{12.51k \times (60 + 1)}]} = 22.28 = 26.96\ dB$$



 # A small difference can be observed between the amplification A_{11} and amplification A_{12}.

If the anode resistance R_a of the first triode (V1) (inverted output) is slightly lower than the R_a of the second triode (V2) (non-inverted output) the amplifications of the inverted and non-inverted outputs can be equalized. In practice this can be realized by using resistors connected in parallel or in series (or by a combined series-parallel connection of resistors) or by using a trimmer potentiometer instead of a single anode resistor of the first triode (V1)(inverted output).

If the anode resistor of the first triode R_{a1} = 56 kΩ (example above) is replaced by a resistor of 51kΩ:

(A_{11} / A_{12} = 24.38 / 22.28 = 1.094, i.e 9.4 %, 56 kΩ – 0.094 x 56 kΩ ≈ 51 kΩ) the amplifications of the inverted and non- inverted outputs become practically equal.

4.8 CATHODYNE PHASE INVERTER (CONCERTINA INVERTER)

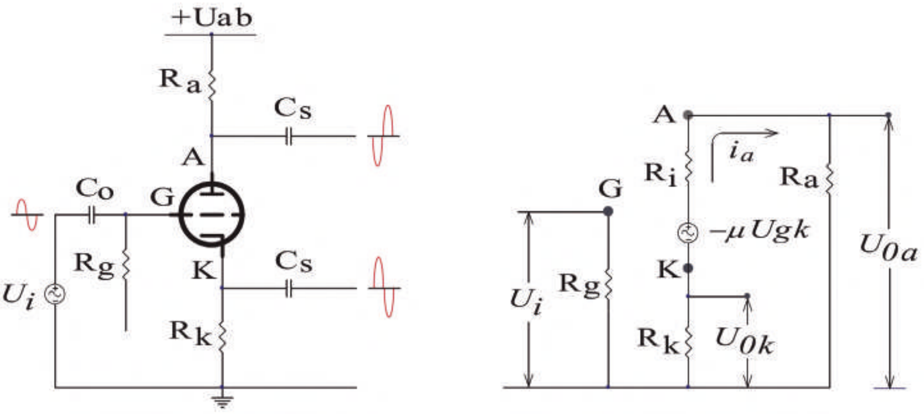


Fig. 4-18

A simplified explanation of Concertina Inverter operation
Assuming a positive signal is applied to the input (grid) of the triode, the current flowing through the triode increases, and the voltage drop across the anode resistor also increases – the anode voltage decreases (inverted output), but, at the same time, the voltage drop across the cathode resistor increases – the cathode voltage increases (non inverted output). Since identical currents flow through anode and cathode resistors, which are of equal resistance, the amplitudes of the signals generated across them are equal, but the signals are opposite in phase.

(A similar process but, with opposite effects, occurs when a negative signal is applied to the input (grid) of the triode).

This type of circuit is commonly used as a driver stage in low-power push-pull output amplifiers.

Amplification and output impedance:

$$U_{gk} = U_i - U_k = U_i - R_k \times i_a$$

Amplification (anode output):

$$A_{0a} = \frac{U_{0a}}{U_i} = -\mu \times \frac{R_a}{R_a + R_i + (\mu+1) \times R_k}$$

Output impedance (anode output):

$$R_{out-a} = R_i' \| R_a = \frac{R_i' \times R_a}{R_i' + R_a} = \frac{R_a \times [R_i + (\mu+1) \times R_k]}{R_a + R_i + (\mu+1) \times R_k}$$

#Note: $R_i' = R_i + (\mu + 1) \times R_k$

Amplification (cathode output): $$A_{0k}=\frac{U_{0k}}{U_i}=\frac{\mu}{\mu+1}\times\frac{R_k}{\frac{R_i+R_a}{\mu+1}+R_k}=\mu\times\frac{R_k}{R_a+R_i+(\mu+1)\times R_k}$$

Output impedance (cathode output): $$R_{out-k}=\frac{\frac{R_i+R_a}{\mu+1}\times R_k}{\frac{R_i+R_a}{\mu+1}+R_k}=\frac{R_k\times(R_i+R_a)}{R_i+R_a+(\mu+1)\times R_k}$$

If **R_a = R_k = R**: $A_{0k}=-A_{0a}=\frac{\mu\times R}{R_i+(\mu+2)\times R}$

 # Note on the output impedance of the cathodyne phase inverter

If we consider not only the change in impedance of the source, but also the change in the voltage generated by the source (under the condition that both outputs are simultaneously loaded with an equal load), both outputs have equal effective output impedance:

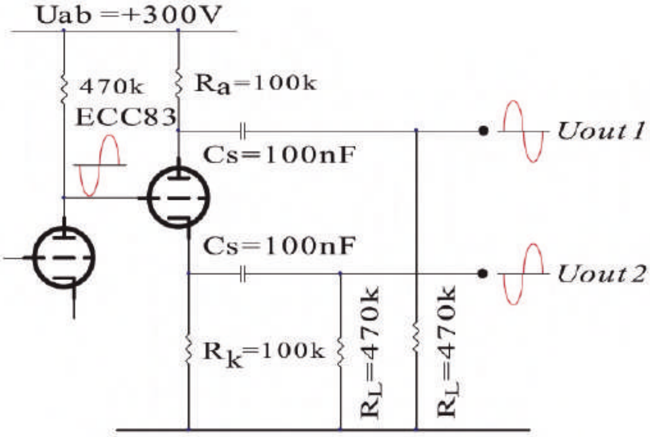
$$R_{out-k}=R_{out-a}=\frac{R\times(R_i+R)}{R_i+(\mu+2)\times R}$$

Main characteristics of Cathodyne phase inverter:

- the **output signal in the anode circuit** and the input signal are **opposite in phase.**
- the **output signal in the cathode circuit** and the input signal are **in phase.**
- low amplification
- good amplitude characteristic
- low harmonic distortion

Example 408.1:

Cathodyne phase splitter.
Tube: ECC83 (12AX7)
μ = 100, R_i = 62.5 kΩ
Power supply: U_{ab} = 300V
R_a = R_k = R = 100 kΩ



$$A_1=-A_2=\frac{\mu\times R}{R_i+(\mu+2)\times R}$$

$$=\frac{100\times 100k}{62.5k+(100+2)\times 100k}=0.974$$

$$R_{out\,1\,(anode)}=\frac{R\times[R_i+(\mu+1)\times R]}{R_i+(\mu+2)\times R}$$

$$=\frac{100k\times[62.5k+(100+1)\times 100k]}{62.5k+(100+2)\times 100k}=99k\Omega$$

$$R_{out\,2\,(cathode)}=\frac{R\times(R_i+R)}{R_i+(\mu+2)\times R}=\frac{100k\times(62.5k+100k)}{62.5k+(100+2)\times 100k}=1.58k\Omega$$

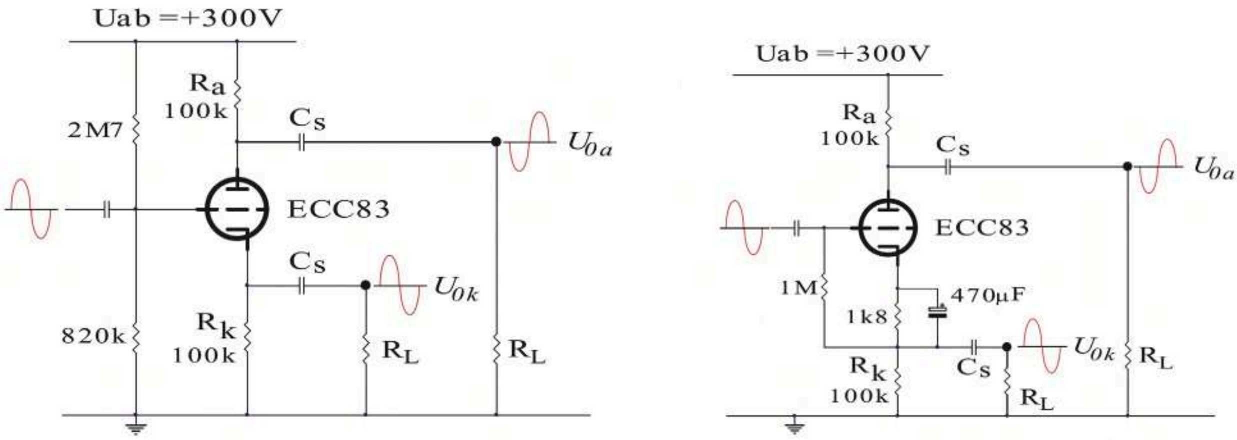


Fig. 4-19

The amplifier circuit explained in Examples 401.3 and 401.3/1 can be redesigned and converted to a cathodyne phase splitter circuit by removing the bypass capacitor of the second triode cathode resistor and using both cathode and anode outputs:

 # Note: The design procedure is identical to the design procedure used in Examples 401.3 and 401.3/1.

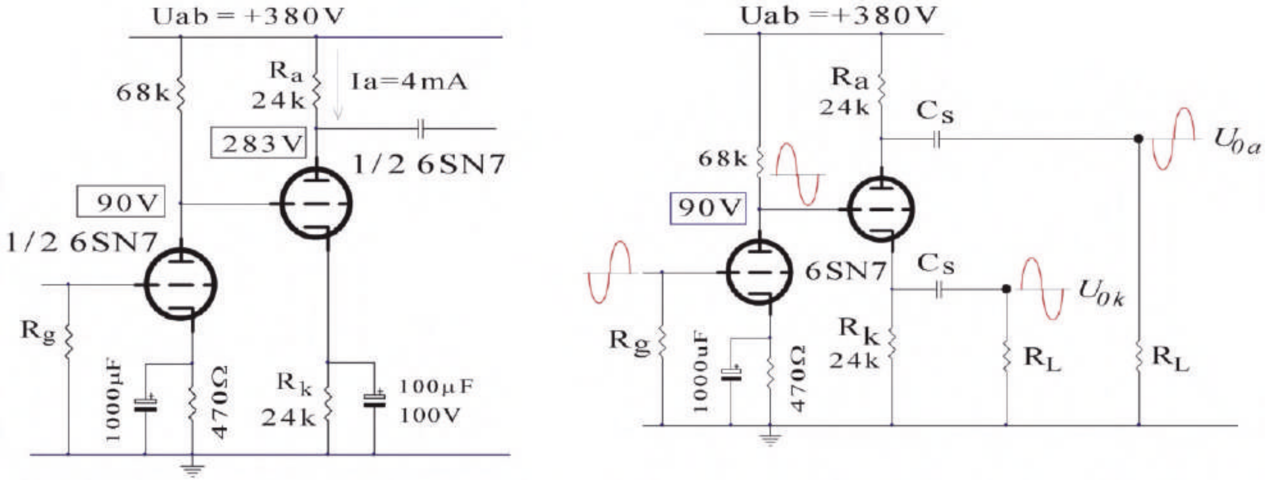


Fig. 4-20

$$A_1(A_{anode}) = A_2(A_{cathode}) = \frac{\mu \times R}{R_i + (\mu + 2) \times R} = \frac{20 \times 24k\Omega}{7.7k\Omega + (20 + 2) \times 24k\Omega} = 0.896$$

$$R_{Oa} = \frac{R \times [R_i + (\mu + 1) \times R]}{R_i + (\mu + 2) \times R} = \frac{24k\Omega \times [7.7k\Omega + (20 + 1) \times 24k\Omega]}{7.7k\Omega + (20 + 2) \times 24k\Omega} = 22.9\ k\Omega$$

$$R_{Ok} = \frac{R \times (R_i + R)}{R_i + (\mu + 2) \times R} = \frac{24k\Omega \times (7.7k\Omega + 24k\Omega)}{7.7k\Omega + (20 + 2) \times 24k\Omega} = 1.42\ k\Omega$$

4.9 MULTI-STAGE AUDIO FREQUENCY AMPLIFIERS

Audio amplifiers should amplify the AC voltage of a certain frequency band as much as possible with minimal signal distortion. Usually, in order to achieve the desired voltage amplification, it is necessary to connect two or more amplifier stages.

The coupling of the amplifier stages can be realized by using a resistance or resistance – capacitance coupling circuit or by using a transformer circuit.

4.9.1 RESISTANCE–CAPACITANCE (RC) COUPLED AMPLIFIER (Fig. 4-21)

The amplified input voltage generated at the anode circuit of the first tube can be amplified again by applying it to the input (grid) of the second tube. The anode circuit of the first tube and the grid of the second tube are coupled with the capacitance C_s in order to prevent the positive anode DC voltage of the first tube from reaching the grid of the second tube and at the same time to conduct the amplified AC signal to the grid of the second tube. The capacitance Cs and R''_g form an AC voltage divider, so the impedance C_s must be small compared to the value of the resistance R''_g (for max. voltage amplification).

The value of R_g is limited by the type of tube. The values of R_g and C_s determine the low end (low cutoff frequency) of the frequency characteristic of the amplifier.

U_g (grid bias voltage) of the tube of the second amplifier stage must fulfill the condition:
minimum $U_g \geq \sqrt{2} \times U_{eff}$; U_{eff} (U_{RMS}) is the voltage (U_1) of the AC signal of the anode circuit of the first tube.

The amplification of the first amplifier stage is: $A_1 = \frac{U_1}{U_{in}}$ $\quad A_1(dB) = 20 \times log\frac{U_1}{U_{in}}$ (409.1)

The amplification of the second amplifier stage is: $A_2 = \frac{U_{out}}{U_1}$ $\quad A_2(dB) = 20 \times log\frac{U_{out}}{U_1}$ (409.2)

The total amplification is: $A = \frac{U_{out}}{U_{in}} = A_1 \times A_2$ $\quad A(dB) = A_1(dB) + A_2(dB)$ (409.3)

In general: $A(dB) = \sum_{i=1}^{n} A_i(dB)$ (409.4)

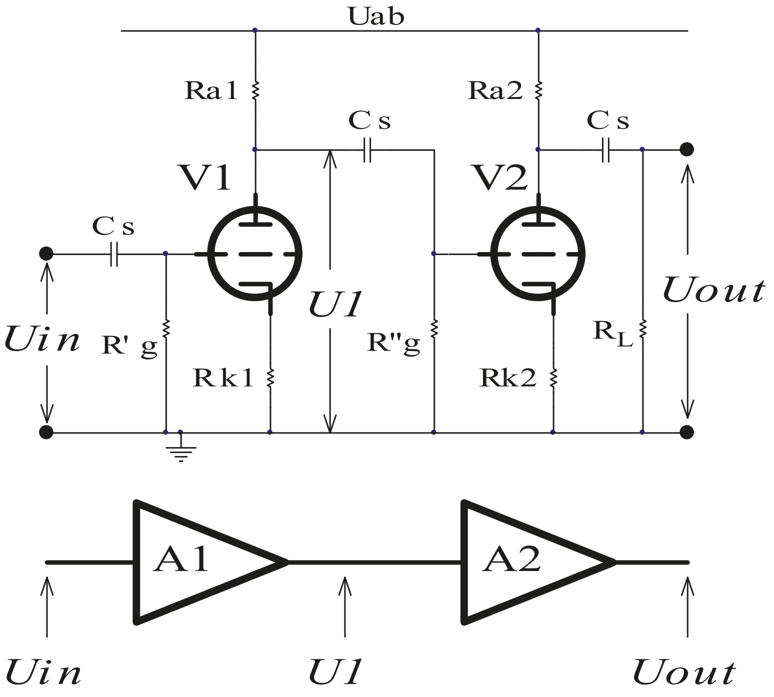
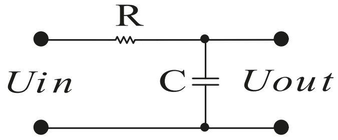


Fig. 4-21

4.9.2 PASSIVE RC FILTERS

- **The low-pass filter** allows the low frequency signals to pass through it unchanged (unchanged amplitudes and phases), but attenuates the high frequency signals and causes their phase lag.



Cutoff Frequency:

$$f_c = \frac{1}{2 \times \pi \times R \times C}$$

Phase Angle:

$$\varphi = -arc\tan(2\times\pi\times f\times R\times C)$$

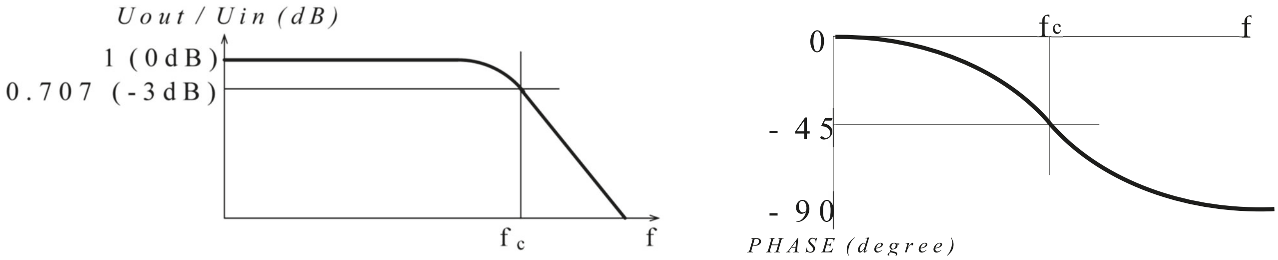
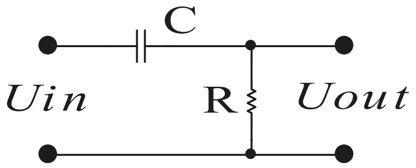


Fig. 4-22

- **The high-pass filter** allows the high frequency signals to pass through it unchanged (unchanged amplitudes and phases), but attenuates the low frequency signals and causes their phase lead.



Cutoff Frequency:

$$f_c = \frac{1}{2\times\pi\times R\times C}$$

Phase Angle:

$$\varphi = arc\tan\frac{1}{(2\times\pi\times f\times R\times C)}$$

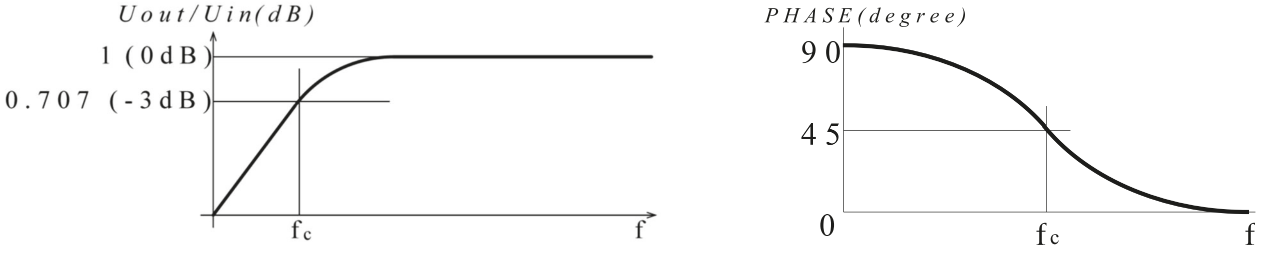


Fig. 4-23

- **The band- pass filter** is a combination of high-pass and low-pass filters. It allows signals of certain frequencies within a certain frequency band to pass through it unchanged, but it attenuates high and low frequency signals at the ends of the frequency band.

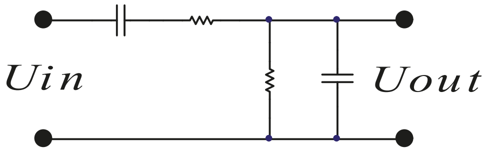


Fig.4-24

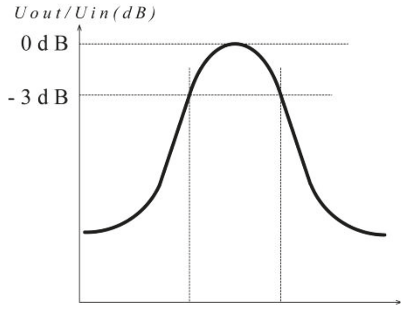


Fig. 4-25

The amplitude characteristic of the amplifier is the frequency band determined by cutoff frequencies, low frequency cutoff and high frequency cutoff, where the amplification is $\frac{1}{\sqrt{2}}$ **(or –3dB)** of the amplification in the middle of the frequency band. Thus, if in the middle of the frequency band the amplification of the amplifier is A, the frequency or **amplitude characteristic** of the amplifier is defined as the frequency band limited by the cutoff frequencies (low (f_{min}) and high (f_{max})) at which the amplification is reduced to $\frac{A}{\sqrt{2}}$ or **–3dB.**

A more detailed analysis of the Resistance – Capacitance Coupled Amplifier shows that there is a **band-pass** filter between the two amplifier stages, see Fig.4-26.

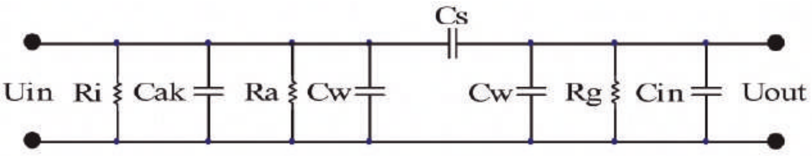


Fig. 4-26

Ri	Internal resistance of the first tube
Cak	Capacitance between anode and cathode
Ra	Anode resistance of the first tube
Cw	Capacitance caused by components wiring
Cs	Coupling capacitor
Rg	Grid resistor of the second tube
Cin	Input capacitance of the second tube

$$f_{min} = \frac{1}{2 \times \pi \times C_S \times \left(R_g + \frac{R_i \times R_a}{R_i + R_a}\right)} \qquad (409.5)$$

$$f_{max} = \frac{1}{2 \times \pi \times C_{eq} \times R_a'} \qquad (409.6)$$

$$\frac{1}{R_a'} = \frac{1}{R_i} + \frac{1}{R_a} + \frac{1}{R_g}; \quad C_{eq} = C_{ak} + C_w + C_w + C_{in}$$

f_{min} – is determined by the characteristic of the tube (R_i), R_a and the coupling capacitor C_s.

f_{max} – is determined by the characteristic of the tube (R_i,C_{ak},C_{gk},C_{ga}), R_a, R_g and the layout and wiring of electronic components.

 # It is necessary to remember as well as employ the Miller effect

$$C_{in} = C_{gk} + C_{ak} \times (\mu + 1) \;\; ; \;\; C_{in} = C_{gk} + C_{ak} \times (A + 1)$$

Example 409.1:

Two stage amplifier:

Tube: ECC83: ($\mu = 100$*;* $R_i = 62.5\ k\Omega$*;* $C_{gk} = 1.6\ pF$*;* $C_{ga} = 1.7\ pF$*;* $C_{ak} = 0.46\ pF$*)*

$R_{a1} = 264\ k\Omega$; $R_{k1} = 3.9\ k\Omega$; $R_{a2} = 100\ k\Omega$; $R_{k2} = 1\ k\Omega$; $R_{g1,2} = R_L = 1\ M\Omega$; $C_s = 0.1\ \mu F$.

Calculate:

1. *Amplification of the first and second amplifier stage and total amplification.*

2. *Amplification at a frequency of 10 kHz.*

3. f_{min} *and* f_{max} *(-3dB).*

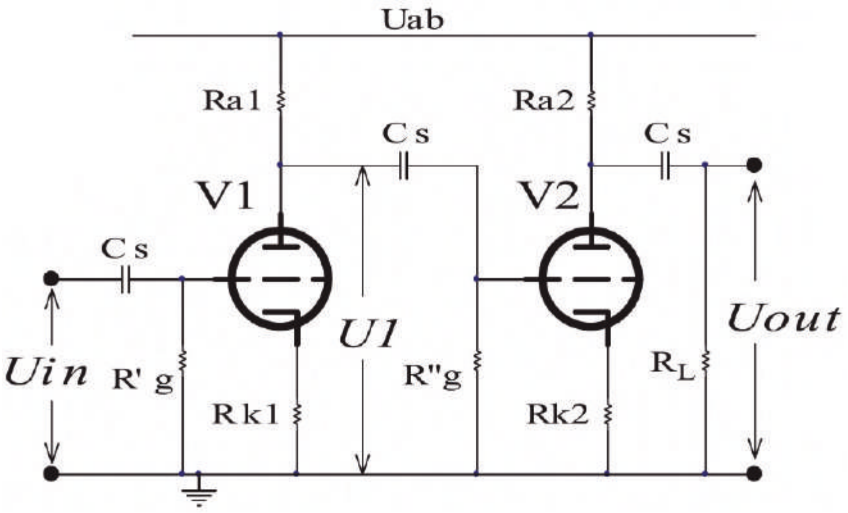


Fig. 4-27

1. Amplification of the first amplifier stage when the second amplifier stage is disconnected:

$$\boldsymbol{R'_{a1}} = R_{a1}\ II\ R_L = 264k\Omega\ II\ 1000k\Omega = \mathbf{208.8}\ \boldsymbol{k\Omega}$$

$$\boldsymbol{A_1} = \mu \times \frac{R'_{a1}}{R_{i1}+R'_{a1}+(\mu+1)\times R_{k1}} = 100 \times \frac{208.8}{62.5+208.8+(100+1)\times 3.9} = \mathbf{31.389} \qquad (\mathbf{29.93}\ \boldsymbol{dB})$$

Amplification of the second stage is:

$$\boldsymbol{R'_{a2}} = R_{a2}\ II\ R_L = 100k\Omega\ II\ 1000k\Omega = \mathbf{90.9}\ \boldsymbol{k\Omega}$$

$$\boldsymbol{A_2} = \mu \times \frac{R'_{a2}}{R_{i1}+R'_{a2}+(\mu+1)\times R_{k2}} = 100 \times \frac{90.9}{62.5+90.9+(100+1)\times 1} = \mathbf{35.73} \qquad (\mathbf{31.06}\ \boldsymbol{dB})$$

The total amplification is:

$$A = A_1 + A_2 = 31.389 \times 35.75 = 1121.5(61\,dB)$$

$$A(dB) = A_1(dB) + A_2(dB) = 29.93 + 31.06 = 61\,dB$$

2. Input capacitance of the second amplifier stage:

$$\boldsymbol{C_{in2}} = C_{gk2} + C_{ga2} \times (1 + A_2) = 1.6 + 1.7 \times (1 + 35.73) = \mathbf{64}\ \boldsymbol{pF}$$

The total capacitance C_{eq} acting in parallel with the anode resistance R_{a1} of the first amplifier stage is:

$$C_{eq} = C_{ak} + C_w + C_w + C_{in} = 0.46 + 5 + 5 + 64 = 74.46\ pF$$

The anode impedance of the first amplifier stage is:

$$\boldsymbol{Z_{a1}} = \frac{R'_{a1} \times \frac{1}{j\omega C_{eq}}}{R'_{a1} + \frac{1}{j\omega C_{eq}}} = \frac{R'_{a1}}{1 + j\omega C_{eq} \times R'_{a1}} = \frac{208.8 \times 10^3}{1 + j2\pi \times 10 \times 10^3 \times 74.46 \times 10^{-12} \times 208.8 \times 10^3} = \frac{\mathbf{208.8 \times 10^3}}{\mathbf{1 + \boldsymbol{j}0.976}}$$

$$= \frac{208.8 \times 10^3 - j203.79 \times 10^3}{1.95} = \mathbf{106.9 \times 10^3 - \boldsymbol{j}104.37 \times 10^3}$$

Amplification of the first amplifier stage when the second amplifier stage is connected (at a frequency of 10 kHz):

$$\boldsymbol{A_1} = \mu \times \frac{Z_{a1}}{R_{i1} + Z_{a1} + (\mu + 1) \times R_{k1}}$$

$$= 100 \times \frac{106.9 \times 10^3 - j104.37 \times 10^3}{62.5 \times 10^3 + 106.9 \times 10^3 - j104.37 \times 10^3 + (100 + 1) \times 3.9 \times 10^3}$$

$$= 100 \times \frac{106.9 - j104.37}{563.6 - j104.37} = 0.177 \times \frac{106.9 - j104.37}{1 - j0.185} = 0.177 \times \frac{126.2 - j84.6}{1.034} = \boldsymbol{21.6 - j14.48}$$

The real part of the amplification is:

$$A_{1r} = \sqrt{21.6^2 + 14.48^2} = 26$$

The amplification is decreased from 31.389 to 26, or from 29.93dB to 28.3dB i.e. amplification is lower by -1.63 dB.

$$\varphi = arctg \frac{14.48}{21.6} = 33.8^0$$

Phase is changed from $\varphi = 180^0$ to $\varphi = (180 - 33.8)^0 = 146.2^0$

Total amplification at 10kHz:

$$\boldsymbol{A = A_{1r} \times A_2 = 26 \times 35.73 = 929} \qquad \boldsymbol{A(dB) = 28.3 + 31.06 = 59.36dB} \quad \boldsymbol{phase\ -33.8^0}$$

3.

$$R''_{a1} = R_{a1}\ II\ [R_{i1} + (\mu + 1) \times R_{k1}] = 264\ II\ [62.5 + (100 + 1) \times 3.6] = 167.25\ k\Omega$$

$$f_{min} = \frac{1}{2 \times \pi \times C_S \times (R_{g2} + R''_{a1})} = \frac{1}{2 \times \pi \times 0.1 \times 10^{-6} \times (1 \times 10^6 + 167.25 \times 10^3)} = \boldsymbol{1.36\ Hz}$$

$$R''_a = R_{a1}\ II\ R_{g2}\ II\ [R_{i1} + (\mu + 1) \times R_{k1}] = 264\ II\ 1000\ II\ [62.5 + (100 + 1) \times 3.9] = 143.28\ k\Omega$$

$$f_{max} = \frac{1}{2 \times \pi \times C_{eq} \times R''_a} = \frac{1}{2 \times \pi \times 74.46 \times 10^{-12} \times 143.28 \times 10^3} = \boldsymbol{14.9\ kHz}$$

4.9.3 FEEDBACK

If fraction of the output voltage of the amplifier is fed back to its input via the appropriate circuit, the main characteristics of the amplifier can be changed.

- **Negative Feedback**: The feedback and input signals are **opposite** in phase.
- **Positive Feedback**: The feedback and input signals are **in phase** (mainly applied in oscillators).

The analysis of the feedback effect on the characteristics of the amplifier can be processed using the schematic illustration shown in Fig.4-28.
The feedback circuit (usually designed with passive electronic components – resistors and capacitors) is inserted between the output and input of the amplifier.

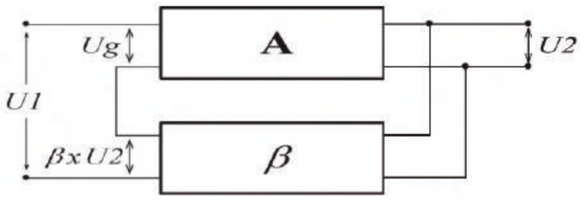


Fig. 4-28

Amplification without feedback:

$$A = \frac{U_2}{U_g} \tag{409.7}$$

If a feedback circuit is applied:

$$\boldsymbol{U_g} = U_1 - \beta \times U_2 \tag{409.8}$$

β – feedback circuit characteristic
β x U_2 – signal at the output of the feedback circuit

$$U_2 = A \times U_1 - A \times \beta \times U_2 \rightarrow U_2 = \frac{A \times U_1}{1 + A \times \beta} \qquad (409.9)$$

Amplification of the amplifier with feedback:

$$A_r = \frac{U_2}{U_1} \qquad (409.10)$$

or:

$$A_r = \frac{A}{1+\beta \times A} \quad \text{or,} \quad A_r = \frac{A}{1-\beta \times A} \qquad (409.11)$$

$F(f) = 1 - \beta \times A$ is the so-called **feedback function**.

If the modulus of the feedback function is greater than 1, the feedback is **negative (negative feedback)**:

$$F(f) = |1 - \beta \times A| > 1 \qquad (409.12)$$

The amplification of an amplifier with negative feedback is lower than the amplification of the same amplifier without feedback.

Why (Negative Feedback)?

- The nonlinear (harmonic) distortion of an amplifier with NFB is lower than the harmonic distortion of the same amplifier without NFB.
- As the amplitude characteristic is almost independent of frequency (if the feedback circuit has constant characteristics in the frequency band), the amplitude distortions are lower - negative feedback flattens frequency response.
- By applying NFB, depending on the type and method of connection, the input impedance of the amplifier can be changed:
  - Feedback circuit connected in parallel with the amplifier input:

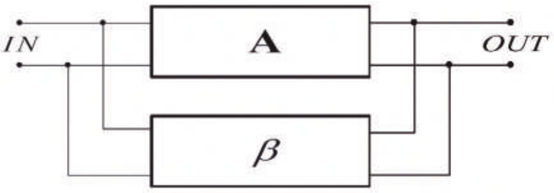


Fig. 429 a

The input impedance of an amplifier with parallel feedback is lower than input impedance of an amplifier without feedback.

$$Z_{in} = \frac{Z_0}{1-\beta \times A} \qquad (409.13)$$

- Feedback circuit connected in series with the amplifier input:

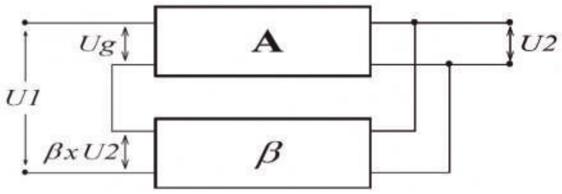


Fig. 4-29 b

The input impedance of the amplifier with serial feedback is higher than input impedance of an amplifier without feedback.

$$Z_{in} = Z_0 \times (1 - \beta \times A) \qquad (409.14)$$

Example 409.2:

a) Designing a multi-stage audio preamplifier by connected the amplifier circuits designed in Examples 401.2, 401.1 (use R_k = 1kΩ) and 402.1. Calculating the total amplification.

b) By applying and calculating NFB, synthesize a multi-stage amplifier with a total amplification of 20dB (with feedback).

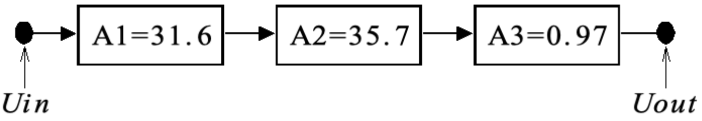


Fig. 4-30

a) Total amplification:

$$\boldsymbol{A} = \frac{U_{out}}{U_{in}} = A_1 \times A_2 \times A_3 = 31.6 \times 35.7 \times 0.97 = \mathbf{1094}$$

- Amplitude characteristic or frequency response (calculation of coupling capacitor C_S):

** Coupling between the first and second tube: RC coupling.*

R_i = 62.5 kΩ, R_a = 264 kΩ, grid resistance of the second tube: R_g = 1MΩ

If the low cutoff frequency is: ***f_{min} = 1Hz****, using Equation 208.9:*

$$R_{a1}^{"} = R_{a1}\ II\ [R_{i1} + (\mu + 1) \times R_{k1}] = 264\ II\ [62.5 + (100 + 1) \times 3.9] = 167.25\ k\Omega$$

$$\boldsymbol{C_S} = \frac{1}{2\times\pi\times f_{min}\times(R_g+R_a^{"})} = \frac{1}{2\times3.14\times1\times(10^6+167.25\times10^3)} = \mathbf{0.136\ \mu F}$$

Standard: ***C_S = 0.22*** μF

Calculation of the high cutoff frequency ***f_{max}*** *:*

$$R_a^{"} = R_{a1}\ II\ R_{g2}\ II\ [R_{i1} + (\mu + 1) \times R_{k1}] = 264\ II\ 1000\ II\ [62.5 + (100 + 1) \times 3.9] = 143.28\ k\Omega$$

The input capacitance C_{in} of the second amplifier stage is (reminder: Miller effect):

$$C_{in} = C_{gk} + C_{ga} \times (A_2 + 1) = 1.6\ pF + 1.7\ pF \times (35.73 + 1) = 64\ pF$$

$$C_{eq} = C_{ak} + C_w + C_w + C_{in} = 0.46\ pF + 5\ pF + 5\ pF + 64\ pF = 74.46\ pF$$

$$f_{max} = \frac{1}{2\times\pi\times C_{eq}\times R_a^{"}} = \frac{1}{2\times\pi\times74.46\times10^{-12}\times143.28\times10^3} = 14.9\ kHz$$

 # Amplitude characteristic – the audio frequency band (especially f_{max}) is extended by applying negative feedback.

b) Amplification with feedback:

$$\boldsymbol{A_r} = \frac{A}{1+\beta\times A} \quad \rightarrow \quad \boldsymbol{\beta} = \frac{A-A_r}{A\times A_r} = \frac{1094-10}{1094\times 10} = \mathbf{0.1}$$

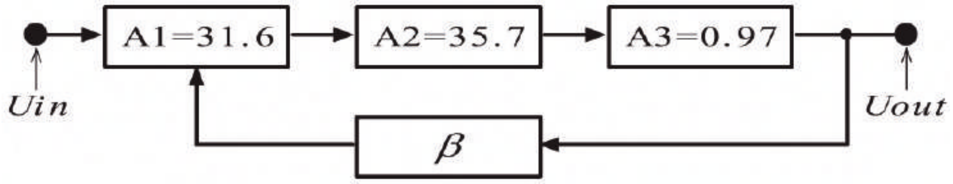


Fig. 4-31

Fig.4-32:

$$\boldsymbol{\beta} = 0.1 = \frac{R_{k1}}{R_{k1}+R_r} \quad \rightarrow \quad \boldsymbol{R_r} = \frac{R_{k1}-0.1\times R_{k1}}{0.1} = \mathbf{35\ k\Omega}$$

Total amplification with negative feedback:

$$\boldsymbol{A_r} = \frac{A}{1+\beta\times A} = \frac{A}{1+\frac{R_{k1}}{R_{k1}+R_r}\times A} = \frac{1094}{1+\frac{3.9}{3.9+35}\times 1094} = \mathbf{9.9} \quad \approx \mathbf{20\ dB}$$

If $R_r = 39\ k\Omega$, $A_r = 10.89$ ≈ 20 dB

Since there is an RC filter at the output of the preamplifier, it is necessary to calculate C_{out1}:

$$\boldsymbol{C_{out1}} = \frac{1}{2\times\pi\times f_{min}\times(R_r+R_{k1})} = \frac{1}{2\times 3.14\times 5\times(39\times 10^3+3.9\times 10^3)} = 0.74\times 10^{-5}\,F = \mathbf{0.74\ \mu F}$$

Standard: ***C_{out1} = 1*** μF (or two 2.2 μF ***connected in series)***

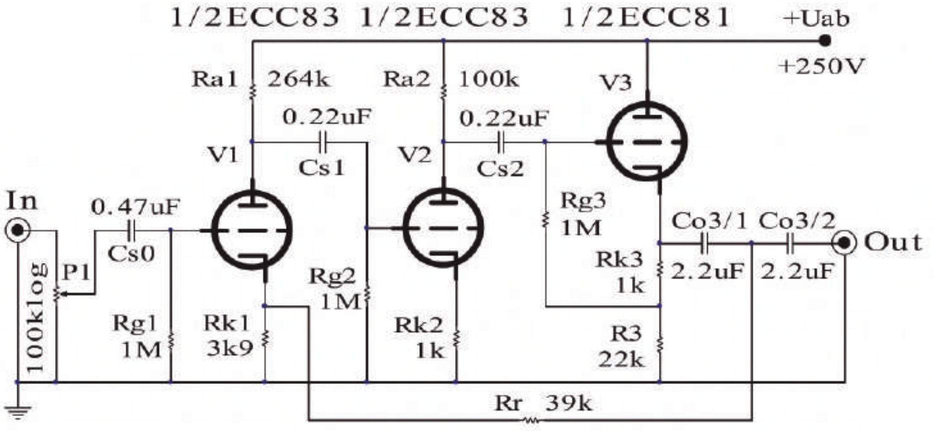
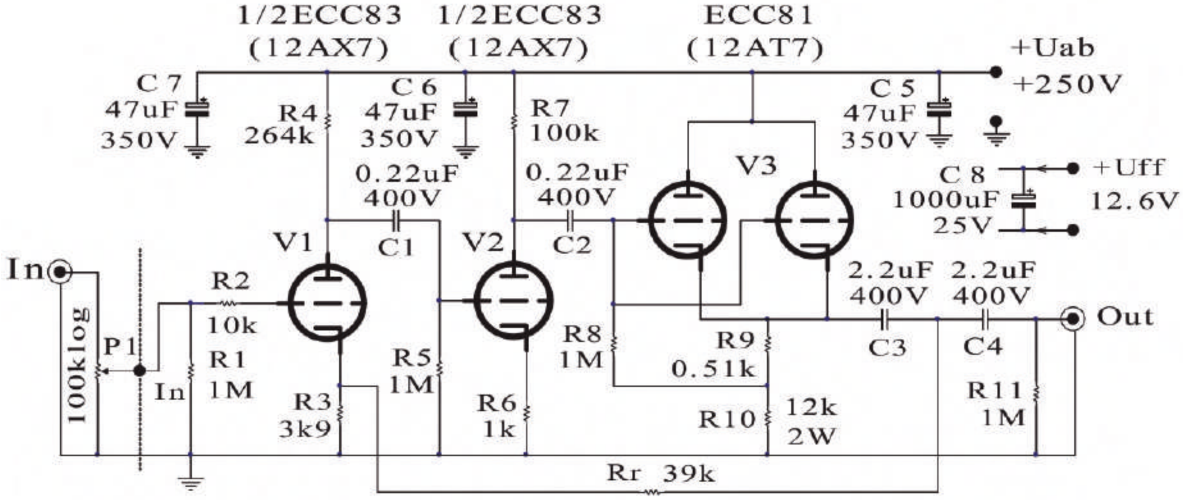


Fig. 4-32

** Characteristics of the preamplifier, Fig. 4-32:*

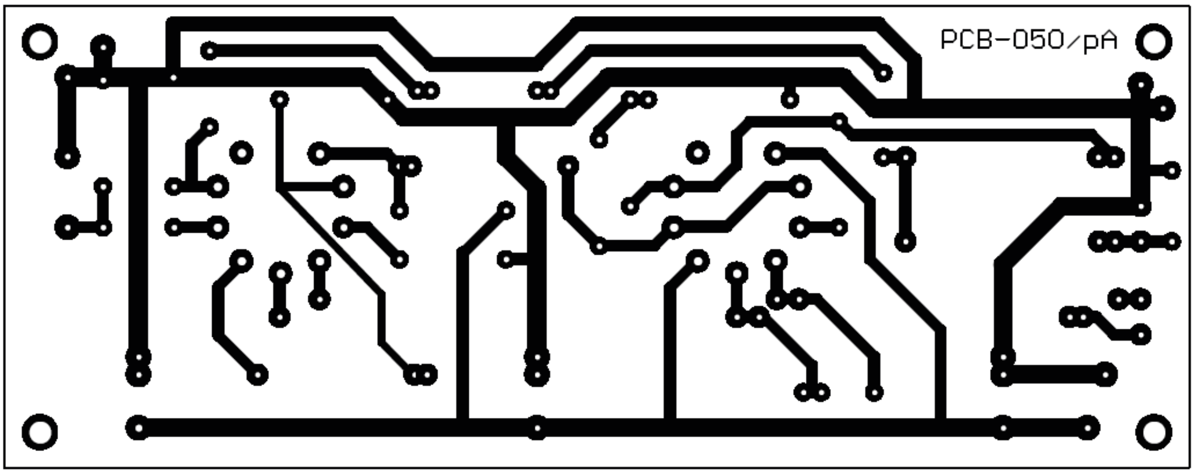
- *very low harmonic distortion (THD ≤ 0.01 %),*
- *low noise (S/N ≥ 80 dB)*
- *low output impedance (a few hundreds of ohms)*
- *good amplitude characteristic (frequency bandwidth: from a few Hz to over one hundred kHz).*

The above example is the basis of typical Hi End preamplifiers used in practice.
The slightly modified version of the line preamplifier explained above can be built using the point – to – point wiring technique and can also be built using the printed circuit board technique.

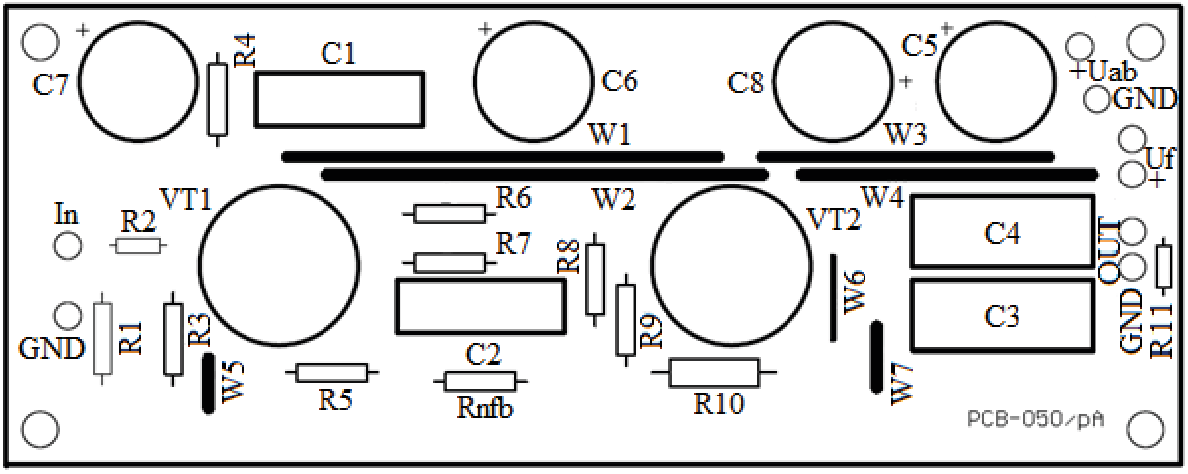


 # Adjust the PCB dimension: (169.5 x 66) mm

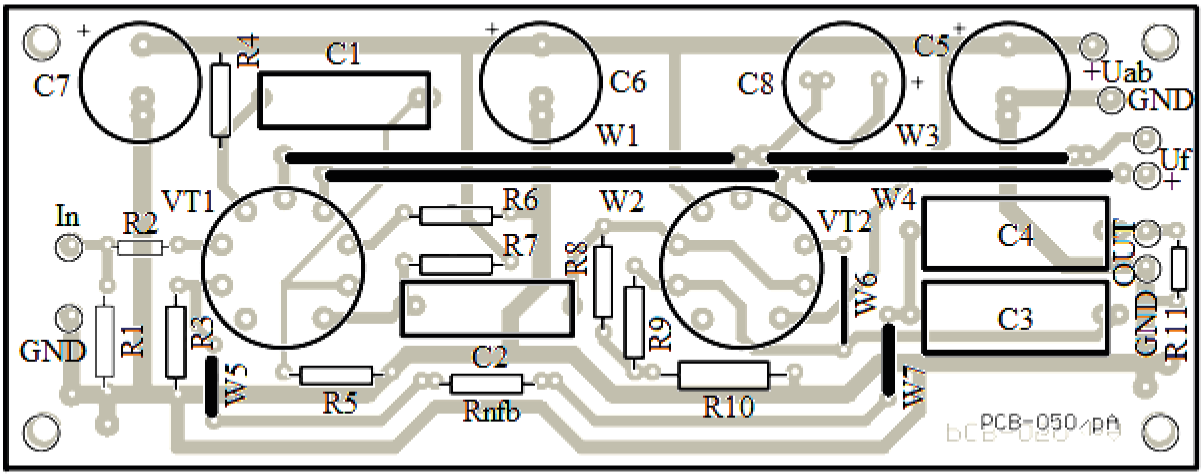
Cu side

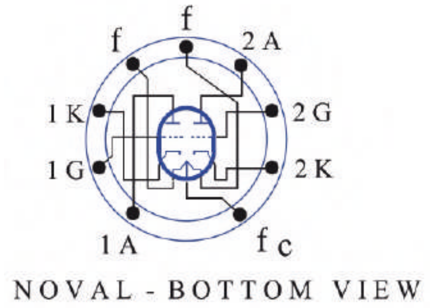


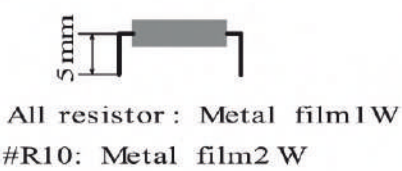
Component side (Component layout)

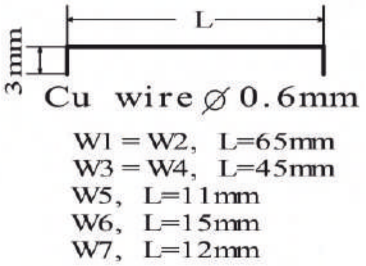


Composite









PCB Laminate: FR 4, tinned Cu side
Drilling: Ø 1.3 mm (tubes socket pins), all others Ø 0.7 mm.

Line amplifier and power amplifier — technical requirements: uniform amplification over a wide frequency range, high input and low output impedance, low noise and low distortion.

However, some audio equipment requires specific characteristics such as well-defined and specific transfer characteristic (amplitude or frequency characteristic (or frequency response)).

A classic example of such equipment is the moving magnet cartridge preamplifier used in a turntable, or the so-called MM RIAA preamplifier.

In these types of preamplifiers, filter networks and frequency-dependent NFB are most often used.

** A technical problem in the recording industry (the master disc cannot accept the full frequency spectrum of the signal at equal amplitude levels) is solved by attenuating the recording signal at frequencies below 1kHz and boosting it over 1kHz in the master-cutting process. This technical issue is standardized by the RIAA (Record Industry Association of America) curve, Fig.4-33 and Table.408.1.*

The recording signal must be inverted in the audio reproduction circuit (preamplifier).

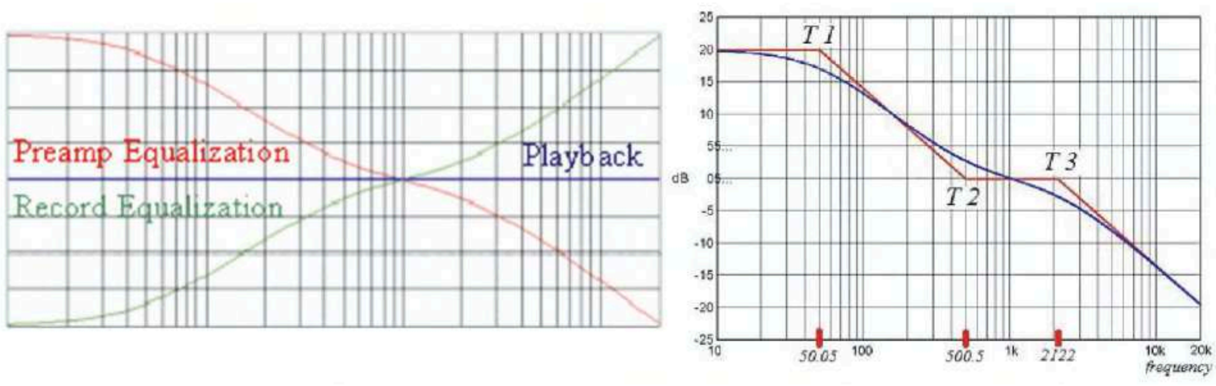


Fig. 4-33

Equalization time constants specified by the RIAA recording / reproduction standard:

3180 µS ***(50.05 Hz)***
318 µS ***(500.5 Hz)***
75 µS ***(2122 Hz)***

The amplitude characteristic increases by 6dB / octave below the frequency of 500Hz, and decreases by 6dB / octave above the frequency of 2122 Hz. Between these two frequencies, the amplitude characteristic is linear and constant.

An equalization curve (amplitude characteristic of a RIAA preamplifier) is exactly the inverse of that stored on a particular record, the combination of the two curves produces an output signal identical to the original.

Table 408.1

FREQUENCY [Hz]	RIAA [dB]
20	+19.27
30	+18.59
40	+17.79
50	+16.95
60	+16.10
70	+15.28
80	+14.51
100	+13.09
125	+11.56
150	+10.27
200	+8.22
250	+6.68
300	+5.48
400	+3.78
500	+2.65
600	+1.84
700	+1.23
800	+0.75
900	+0.35
1000	0.00
1500	-1.40
2000	-2.59
3000	-4.74
4000	-6.61
5000	-8.21

6000	-9.60
7000	-10.82
8000	-11.89
9000	-12.86
10000	-13.73
11000	-14.53
12000	-15.26
13000	-15.94
14000	-16.57
15000	-17.16
16000	-17.71
17000	-18.23
18000	--18.72
19000	-19.18
20000	-19.62

- Passive filter RIAA Equalizer: R_{f1}, R_{f2}, C_{f1}, C_{f2}

- The RIAA equalization can be realized passively (passive RC network in the form of a filter) or actively using frequency-selective feedback of the preamplifier.

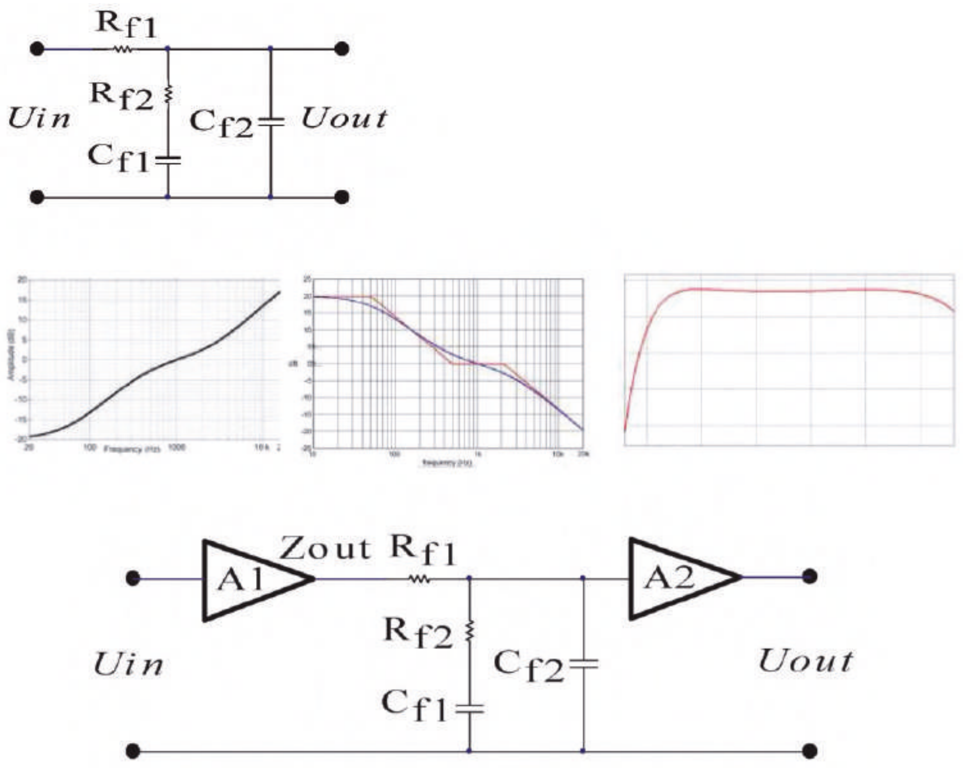


Fig. 4-34

Basic equations used in the RIAA equalizer design process:

a) $r_1 = Z_{out} + R_{f1}$, ($Z_{out} = R_a \text{ II } R_i$) $r_1' = r_1 \text{ II } Z_{in}''$

b) $r_1' \times C_{f1} = 2187\ \mu S$, $r_1' \times C_{f2}' = 750\ \mu S$,
$C_{f2}' = C_{f2} + (C_{in2} + C_{Muller} + C_{stray})$
$R_{f2} \times C_{f1} = 318\ \mu S$

c) $C_{f1} / C_{f2}' = 2.916$, $r_1' / R_{f2} = 6.8774 \div 7.7$

Z_{out} – output impedance of the previous amplifier circuit (if the previous amplifier stage is an anode follower: $Z_{out} = R_i$ II R_a)

Z″ – input impedance of the next amplifier circuit (if the next amplifier stage is an anode follower: Z″ = R_g)

($C_{in2} + C_{Miller} + C_{stray}$) – total input capacitance of the next amplifier stage

C_{Muller} – Miller capacitance
C_{in} – input capacitance of the next amplifier stage
C_{stray} – parasitic capacitance caused by component wiring

Examples:

The modified line preamplifier designed above can be transformed into a MM RIAA preamplifier by inserting an RIAA equalizer circuit (passive filter) between the first and second amplifier stages of the line preamplifier.

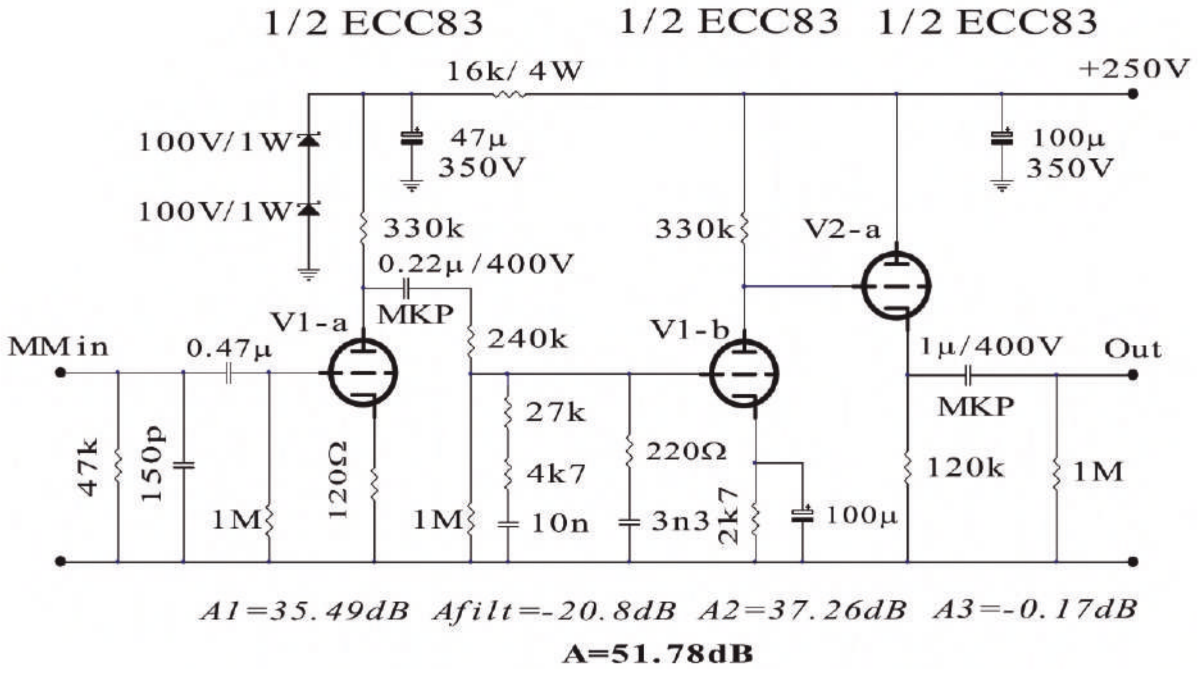


Fig. 4-35

The amplification of the first preamplifier stage is approximately 60 (35.5 dB) and the output resistance is approximately 59 kΩ.

The amplification of the second preamplifier stage is approximately 72.9 (37.26 dB)
The amplification of the third preamplifier stage is 0.97 (- 0.265 dB)
Total amplification without RIAA equalizer circuit is 4215 (72.49 dB)

RIAA equalizer circuit design:
The calculation should start using standard values of electronic components, for example: $\mathbf{C_{f1} = 10\ nF.}$

Using the equation: $R_{f2} \times C_{f1} = 318\mu S \rightarrow \mathbf{R_{f2}} = 318 \times 10^{-6} / 10 \times 10^{-9} = \mathbf{31.8\ k\Omega}$
Two standard resistors **27 kΩ** *and* **4.7 kΩ** *connected in series: (***31.7 kΩ***) ≈ 31.8 kΩ*

Using the equation: $r_1' \times C_{f\,1} = 2187\ \mu S \rightarrow \mathbf{r_1' = 218.7\ k\Omega}$
$r_1' = (R_{out\ I\ stage} + R_{f1})$ II $R_{g\ II\ stage} \rightarrow 218.7\ k\Omega = (59\ k\Omega + R_{f1})$ II $1000\ k\Omega \rightarrow \mathbf{R_{f1} = 221\ k\Omega}$
Using the equation: $C_{f1} / C_{f2}' = 2.916 \rightarrow \mathbf{C_{f2}'} = 10\ nF / 2.916 = \mathbf{3.429\ nF}$

Since C_{f2}' *is:* $C_{f2}' = C_{f2} + (C_{in2} + C_{Miller} + C_{stray}) \rightarrow \mathbf{C_{f2}} = 3.429\ nF – 74.5\ pF = \mathbf{3.35\ nF}$

Standard: $\mathbf{C_{f2} = 3.3\ nF}$

At a frequency of f = 1 kHz, the impedance of the RIAA equalizer circuit, without R_{f1}, is approximately 23.4 kΩ.

*RIAA equalizer attenuation is 23.4 kΩ / (23.4 kΩ + R_{f1}) → **RIAA EQ Attenuation = 0.096 = - 20.37 dB***

***The total amplification of the MM RIAA preamplifier is**:*
Sum of total amplification without RIAA equalizer circuit in dB (72.49 dB) and RIAA EQ Attenuation in dB (-20.37dB):

*72.49 dB – 20.37 dB = **52 dB; (x 398.1)***
An example of a RIAA preamplifier using SRPP amplifier circuits

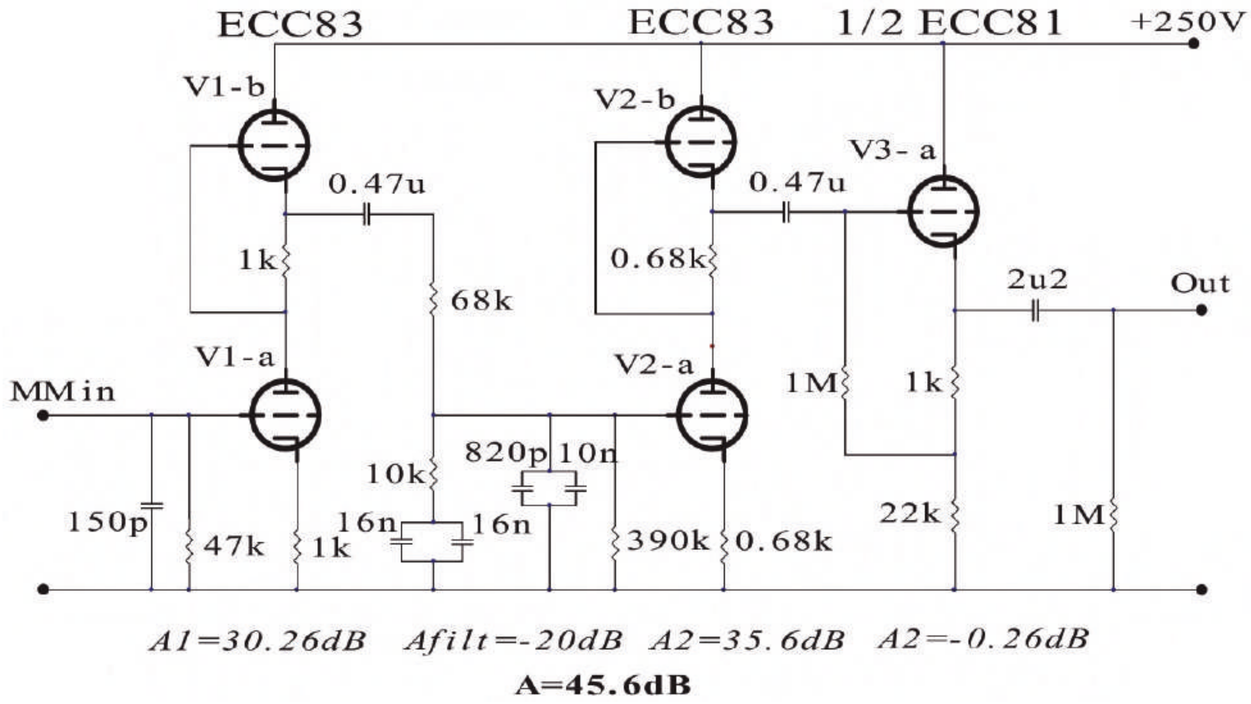
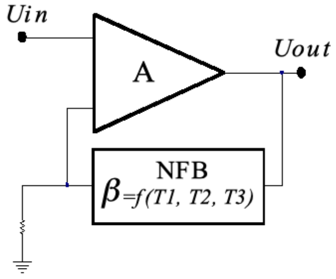


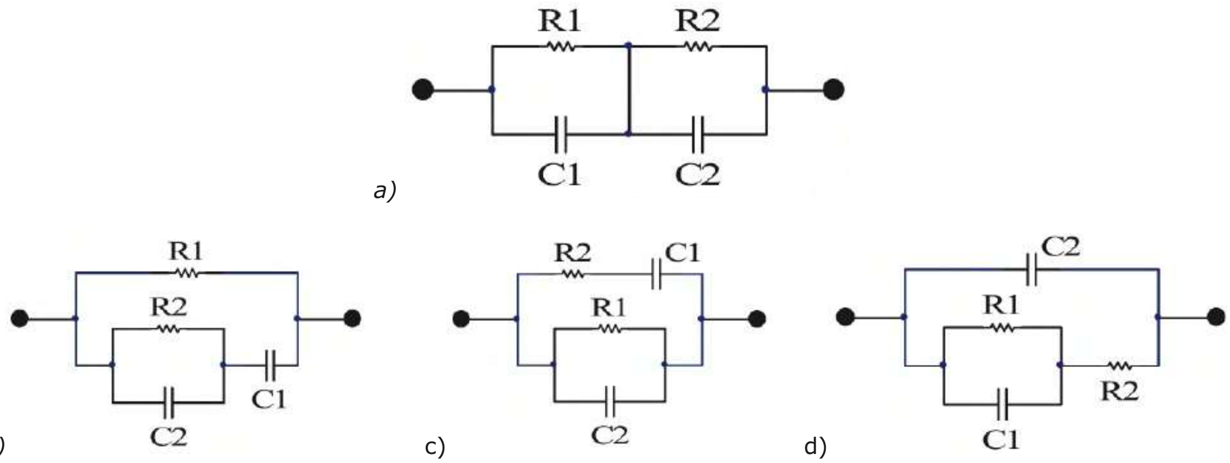
Fig. 4-36

ACTIVE RIAA EQUALIZER

RIAA equalization can be built using a frequency-selective feedback circuit of the preamplifier:



The most commonly used equalization networks (applied in the negative feedback circuit (NFB)):



a)

b) c) d)

Fig. 4-37

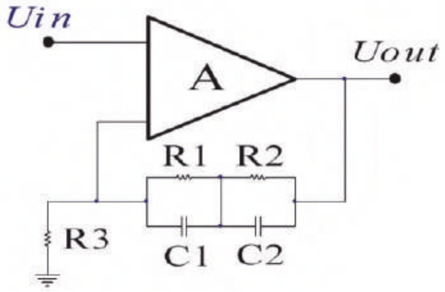


Fig. 4-38

Basic equations - type a) RIAA equalizer circuit:
$T_1 = 3180$ μs; $T_2 = 318$ μs; $T_3 = 75$ μs
$C_1 = 3.6 \times C_2$ $R_1 = 11.7 \times R_2$
$T_1 = R_1 \times C_1$ $T_3 = R_2 \times C_2$
$R_2 \times C_1 = 270$ μs

$$T_2 = (C_1 + C_2) \times \frac{R_1 \times R_2}{R_1 + R_2}$$

NFB network impedance: $Z_{NFB}(j\omega) = (R_1 \text{ II } j\omega C_1) + (R_2 \text{ II } j\omega C_2)$; $\omega = 2\pi f$

$$Z_{NFB}(j\omega) = \frac{R_1}{j\omega C_1 R_1 + 1} + \frac{R_2}{j\omega C_2 R_2 + 1}, \quad \textit{absolute value: } |Z_{NFB}(j\omega)| = \sqrt{Z_{NFB}(j\omega) \times Z_{NFB}(-j\omega)}$$

$$|Z_{NFB}(j\omega)| = \sqrt{\frac{R_1^2(\omega^2 C_2^2 R_2^2 + 1) + 2R_1 R_2(\omega^2 C_1 R_1 C_2 R_2 + 1) + R_2^2(\omega^2 C_1^2 R_1^2 + 1)}{(\omega^2 C_1^2 R_1^2 + 1) \times (\omega^2 C_2^2 R_2^2 + 1)}}$$

Amplification (f = 1 kHz): $A_{1kHz} = \frac{A_0}{1 + \beta \times A_0} = \frac{A_0}{1 + \frac{R_3}{|Z_{NFB}(1kHz)|} \times A_0} \approx \frac{|Z_{NFB}(1kHz)|}{R_3} + 1$

A_0 = amplification without NFB

Example:

Modification of the line preamplifier designed above using a RIAA equalizer network (type a) in the NFB circuit:

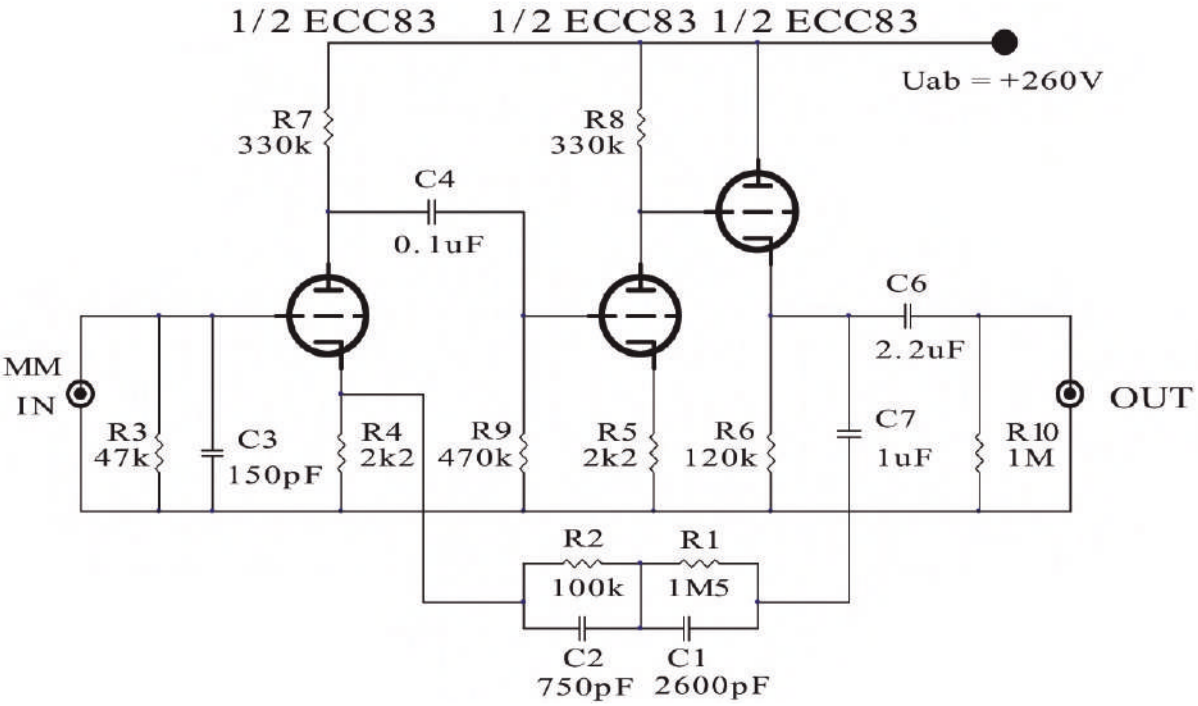


Fig. 4-39

Design and calculation of RIAA equalizer circuit:

The calculation is usually started using standard electronic components: for example ***R_2 = 100 kΩ.***

Using the equation: $T_3 = 75\mu s = R_2 \times C_2$ → ***C_2*** $= 75 \times 10^{-6} / 100 \times 10^3 =$ ***750 pF***

Using the equation: $C_1 = 3.6 \times C_2$ → ***C_1*** $= 3.6 \times 750$ pF $=$ ***2700 pF***

Using the equation: $T_1 = 3180\ \mu s = R_1 \times C_1$ → ***R_1*** $= 3180 \times 10^{-6} / 2700 \times 10^{-12} =$ ***1.178 MΩ***

(In practice, standard electronic components are used: $R_1 = 1.5$ MΩ, $C_1 = 2.6$ nF, $R_2 = 100$ kΩ, $C_2 = 750$ pF*)*

At f = 1 kHz the impedance of the RIAA equalizer circuit is approximately 126.82 kΩ.

The total amplification of the MM RIAA amplifier is:

$$A_{1kHz} = \frac{|Z_{NFB}(1kHz)|}{R_3} + 1 = \frac{126.82\ k\Omega}{2.2\ k\Omega} + 1 = 58.64 = 35.36\ dB$$

A useful tool that can be used in the RIAA preamplifiers testing procedure can be made using an inverse RIAA equalizer circuit based on the Jung – Lipshitz model:

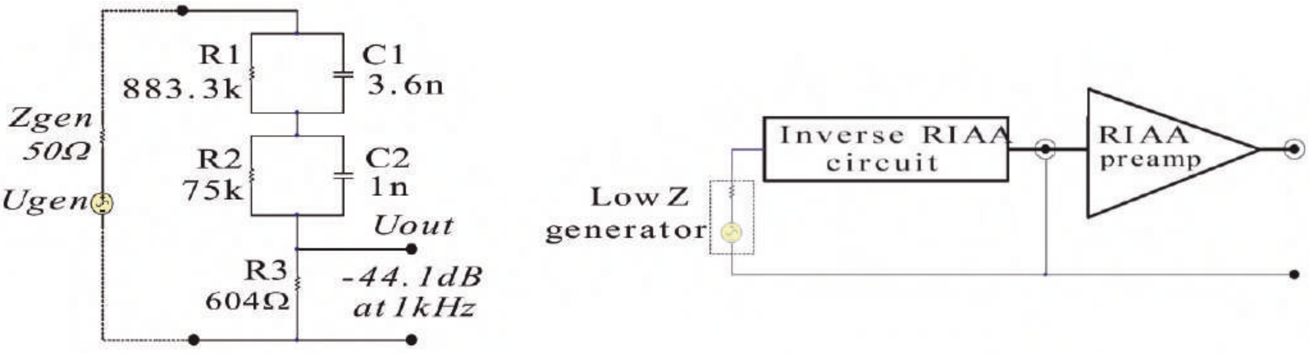


Fig. 4-40

A high-accuracy inverse RIAA circuit can be built by combining standard value resistors:

Fig. 4-41

Example:
Hi End moving magnet cartridge RIAA preamplifier designed using EC900 and 6HM5 tubes manufactured by Ei Nis Radio Tube Factory, Serbia.

Heater Characteristics and Ratings:

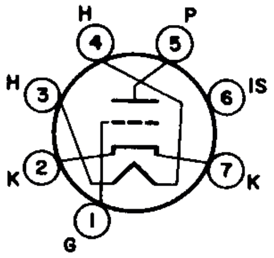
Voltage (AC or DC).	6.3 ± 0.6	volts
Current at heater volts = 6.3	0.180	amp

Characteristics, Class A_1 Amplifier:

Plate Voltage	135	volts
Grid Voltage.	–1	volt
Amplification Factor.	72	
Plate Resistance (Approx.).	5000	ohms
Transconductance.	14500	μmhos

Basing Designation for BOTTOM VIEW. 7GM

Pin 1 – Grid
Pin 2 – Cathode
Pin 3 – Heater
Pin 4 – Heater



Pin 5 – Plate
Pin 6 – Internal Shield
Pin 7 – Cathode

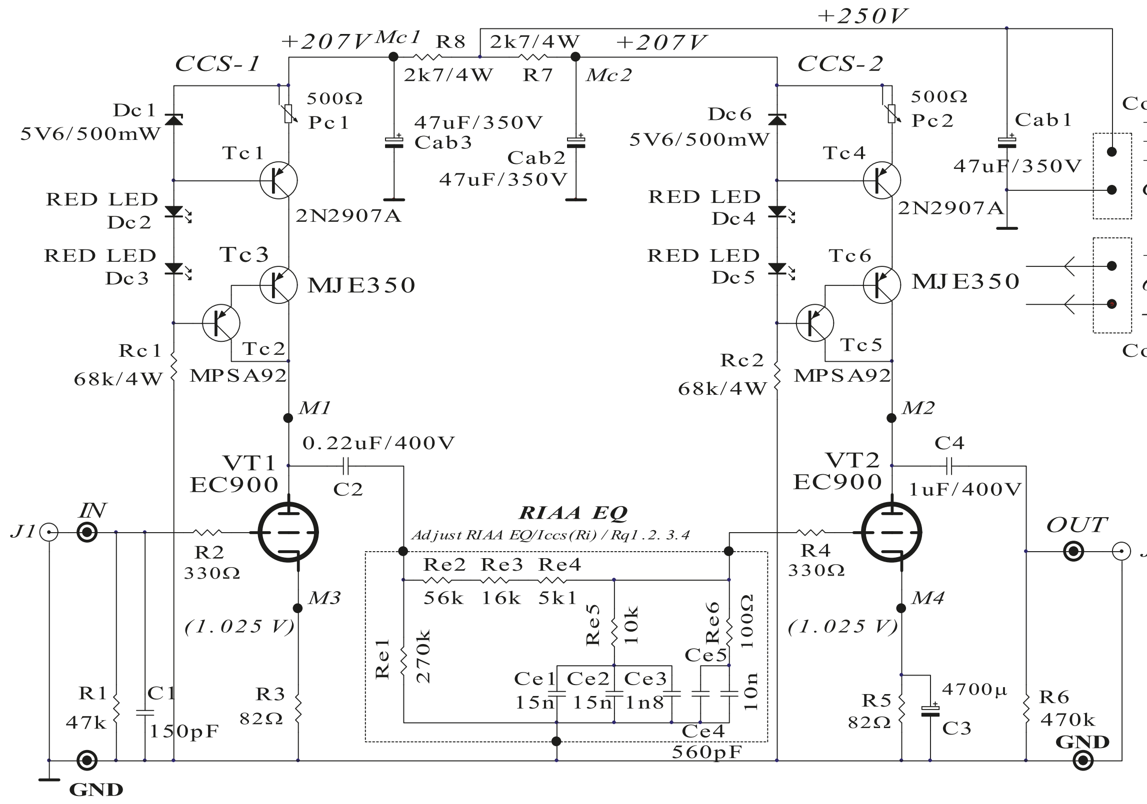
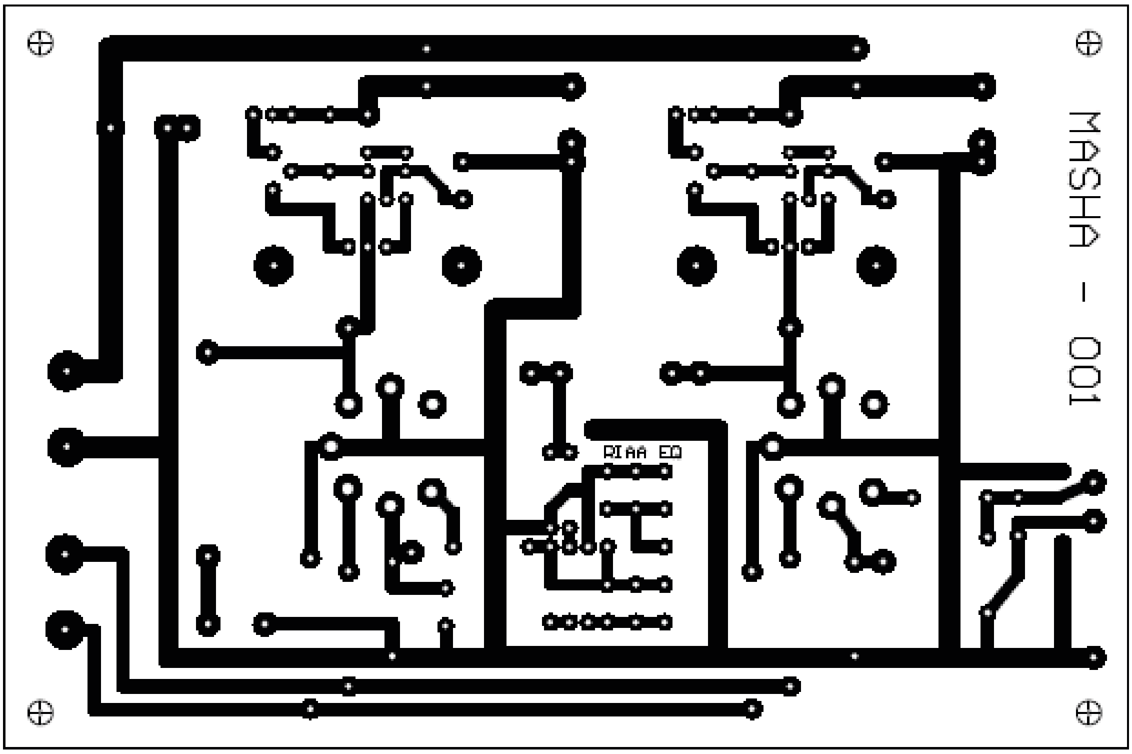


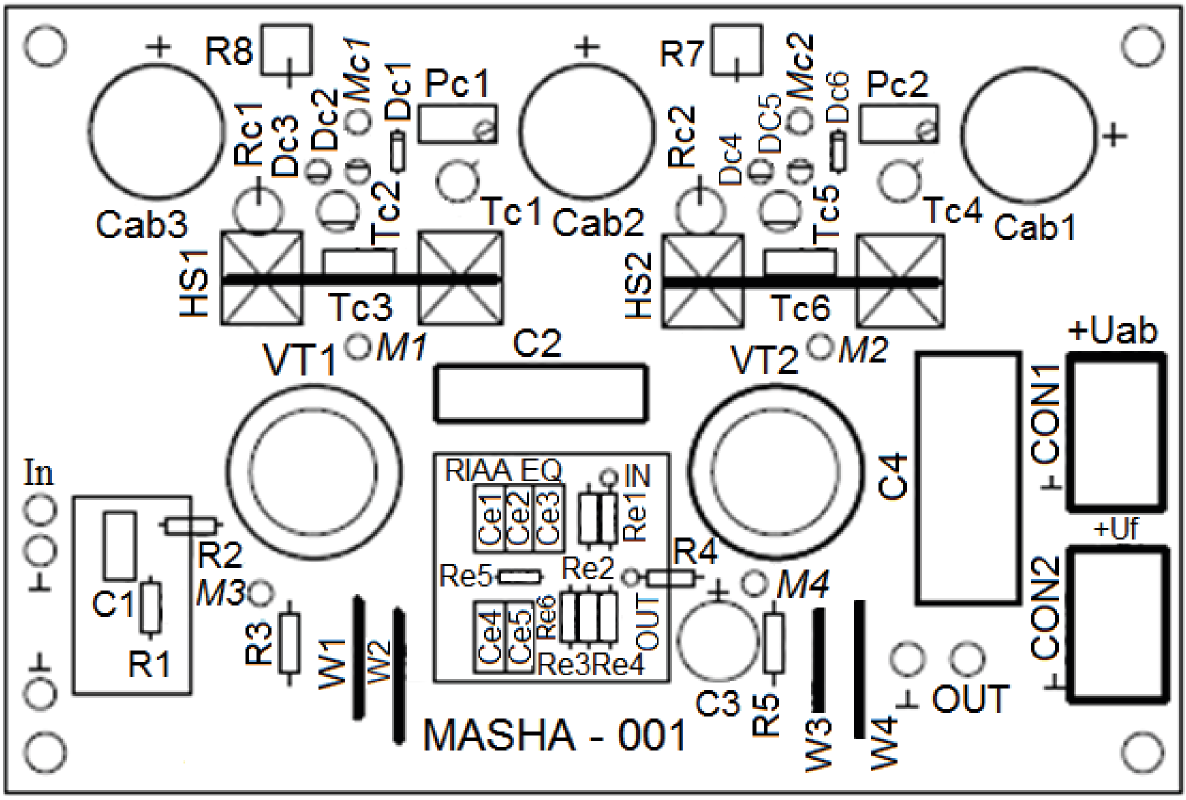
Fig. 4-42

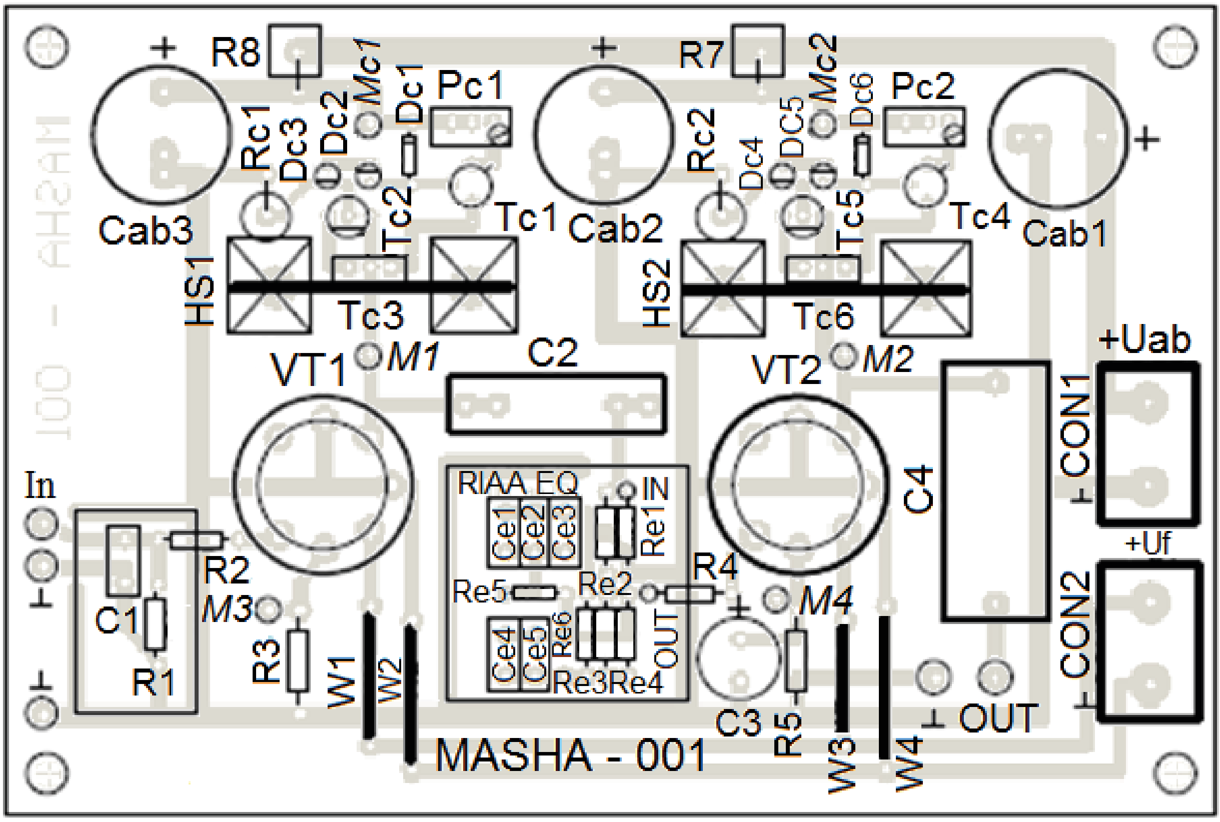
Amplification: 51.78 dB (52.28 dB with C3)
THD: ≤ 0.02%

RIAA error: ≤ ± 0.1dB (20Hz ÷ 20kHz)
S / N ≥ 72 dB



Cu side, PCB dimensions: (150 x 100) mm


R8
Mc1
Dc1
Pc1
Rc1
Dc3
Dc2
Cab3
Tc2
Tc1
HS1
Tc3
R7
Mc2
Dc6
Pc2
Rc2
Dc4
DC5
Tc5
Tc4
Cab2
Cab1
HS2
Tc6
VT1
M1
C2
VT2
M2
+Uab
CON1
+Uf
CON2
C4
In
R2
M3
C1
R1
R3
W1
W2
RIAA EQ
IN
Ce1
Ce2
Ce3
Re1
Re5
Re2
R4
M4
Ce4
Ce5
Re6
OUT
Re3
Re4
C3
R5
W3
W4
OUT
MASHA - 001


R8
Mc1
Dc1
Pc1
Rc1
Dc3
Dc2
Cab3
Tc2
Tc1
HS1
Tc3
R7
Mc2
Dc6
Pc2
Rc2
Dc4
DC5
Tc5
Tc4
Cab2
Cab1
HS2
Tc6
VT1
M1
C2
VT2
M2
+Uab
CON1
+Uf
CON2
C4
In
R2
M3
C1
R1
R3
W1
W2
RIAA EQ
IN
Ce1
Ce2
Ce3
Re1
Re5
Re2
R4
M4
Ce4
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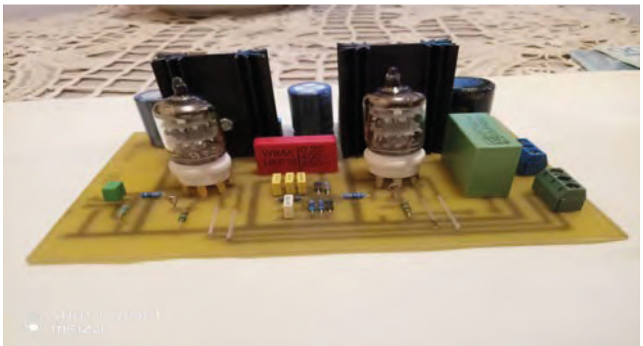
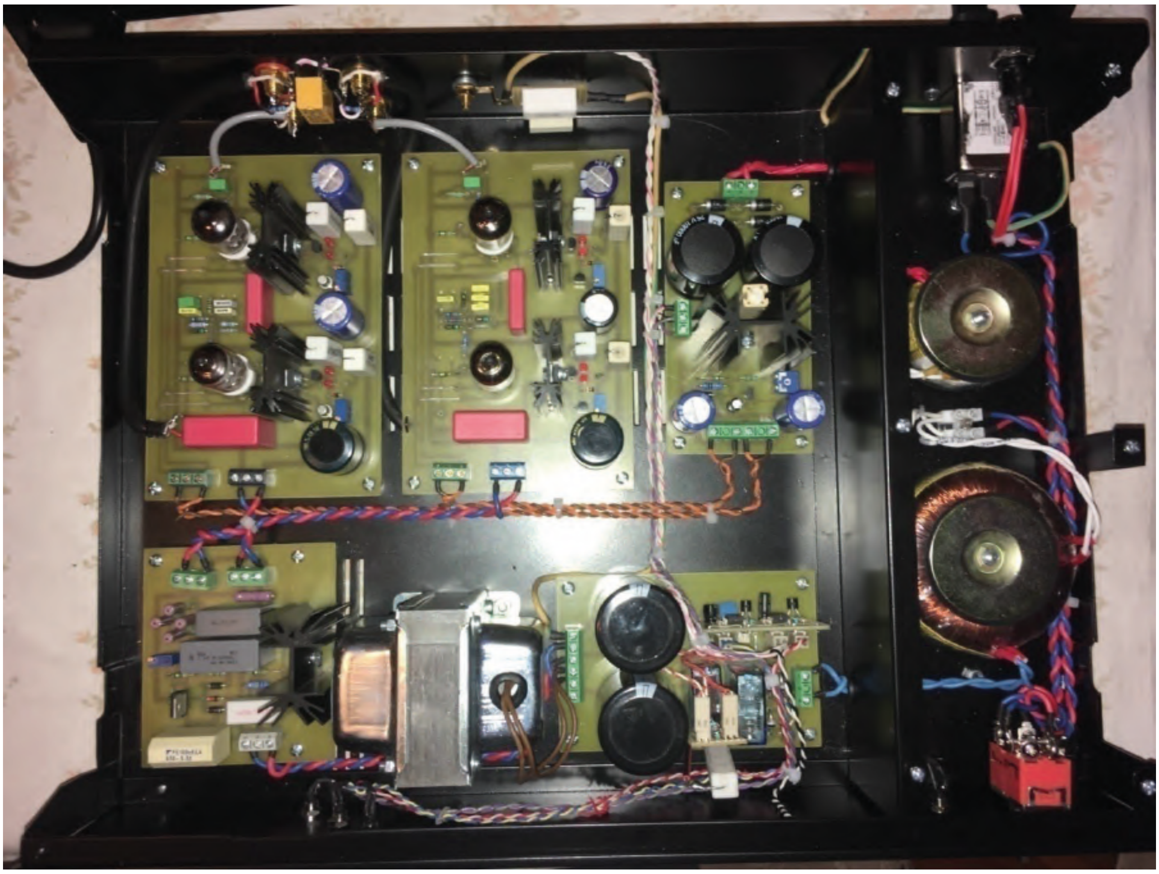
OUT
MASHA - 001


SHOT ON M6T
MEIZU






MM (RIAA) Preamplifier
RUN
VACUUM TUBE
H.V.
Duke Lab
HEAT
M & D
WARMING UP
IN PROGRESS
ON
OFF

4.10 SINGLE-ENDED OUTPUT AMPLIFIER CONFIGURATION

Basically, a single ended output amplifier is a grounded cathode amplifier with a transformer as the anode load. The secondary load of the transformer is the speaker or the speaker box or in other words the speaker as the final load of the amplifier connected to the anode circuit via the transformer (Fig. 4-43).

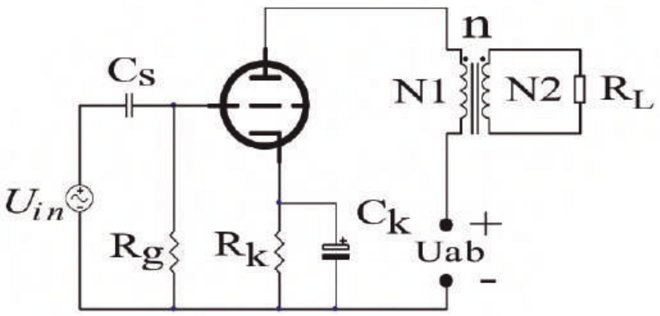


Fig. 4-43

A few facts should be considered:

The anode load (R_a) is a transformer, so

$$R_a = \left(\frac{N_1}{N_2}\right)^2 \times R_L = \frac{R_L}{n^2}$$

$$n = \frac{N_2}{N_1} = \frac{Secondary\ turn\ number}{Primary\ turn\ number} = turns\ ratio$$

R_L – secondary load

The role of the transformer is to adjust the load impedance (speaker) to the optimal anode load by changing the turns ratio and at the same time to prevent the flow of DC anode current through the load (speaker).

If the DC resistance (ohmic resistance) of the transformer primary winding is neglected, the anode voltage is equal to the power supply voltage.

To avoid confusion: As the transformer stores energy in a magnetic field, the anode voltage can exceed the supply voltage – ideal tube: Ua can swing from 0V to 2 x Uab.

If a single ended output amplifier is treated as a classical grounded cathode amplifier, the output power is:

$\boldsymbol{P_{out}} = \frac{1}{2} \times R_a \times I_{am}^2 = \frac{1}{2} \times \left(\frac{\mu \times U_{gm}}{R_i + R_a}\right)^2 \times R_a$; U_{gm} is the amplitude of AC voltage applied to the grid of the tube and I_{am} is the amplitude of the anode current. The anode load R_a for maximum power can be calculated if the first derivative of P_{out} is equal to zero: $\frac{dP_{out}}{dR_a} = 0$ → $\mathbf{R_a = R_i}$ and $\boldsymbol{P_{out\ max}} = \frac{1}{8} \times \frac{\mu^2 \times U_{gm}^2}{R_i}$.

The above analysis is not suitable for practical application. The main and first reason: the tube is an electronic component of limited power – the operating area of the tube is limited by its maximum dissipation of the anode. The second reason is: the tube is a non-linear electronic component.

Graphical analysis of the single ended amplifier implemented in Ia = f(Ua) for Ug = ctc anode characteristics diagram of an ideal tube – straight and equidistant characteristics with a hyperbola of maximum anode dissipation plotted, can help to better clarify the problems related to the optimal anode load necessary to produce the maximum output power of a single ended amplifier.

Graphical analysis:

In order to achieve maximum efficiency of the tube, the bias voltage (grid 1 voltage) must be chosen in such a way that at a given anode voltage the operating point (quiescent point) is placed on the power hyperbola curve (maximum anode dissipation) as shown in Fig.4-44.

The operating area of the tube is limited by points A and B placed on the working line (anode load line) symmetrically with respect to point M (quiescent point). Point A is located on the characteristic of the grid voltage of 0 V, and point B is located on the characteristic of the grid voltage of the anode current cutoff.

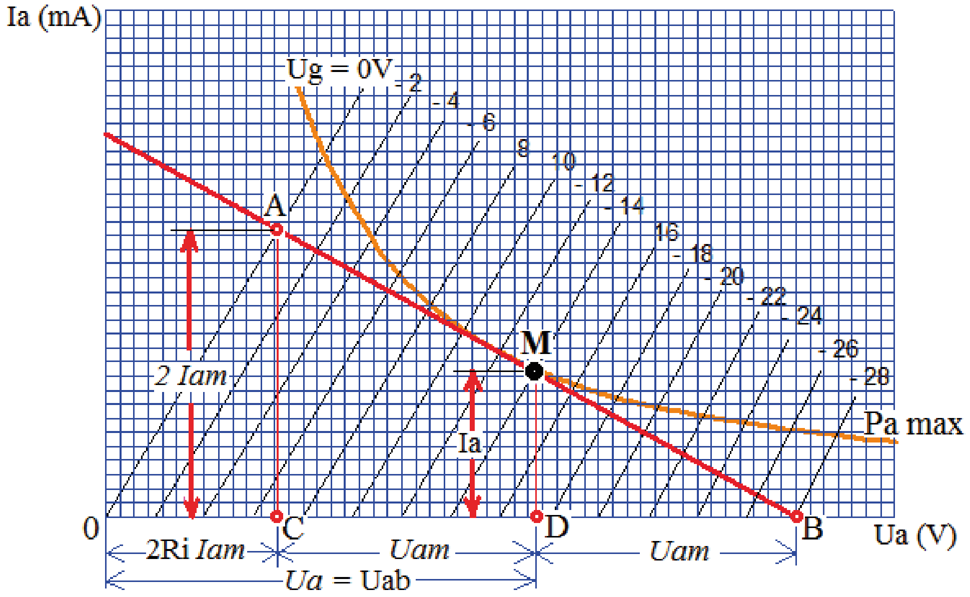


Fig. 4-44

In Fig. 4-44, the amplitude of the AC anode current is equal to the DC anode current:

($I_{am} = I_a$), $U_{ab} = 2 \times R_i \times I_{am} + U_{am}$ and $U_{am} = R_a \times I_{am}$.

By combining the last two equations: $U_{ab} = 2 \times R_i \times I_{am} + R_a \times I_{am}$ *and* $I_{am} = \frac{U_{ab}}{R_a+2\times R_i}$.

Useful output power is: $\boldsymbol{P_{out}} = \frac{1}{2} \times R_a \times I_{am}^2 = \frac{1}{2} \times \frac{U_{ab}^2}{2\times(R_a+2\times R_i)^2} \times R_a$

The anode load R_a for maximum power can be calculated if the first derivative of P_{out} is equal to zero $\frac{dP_{out}}{dR_a} = 0$:

$$\mathbf{R_a = 2 \times R_i} \quad \text{and} \quad \boldsymbol{P_{out\ max}} = \frac{U_{ab}^2}{16\times R_i} .$$

The maximum output power is obtained when the amplitude of the signal applied to the grid of the tube is:

$$\boldsymbol{U_{gm}} = \frac{1}{\mu} \times \frac{R_i+R_a}{R_a+2\times R_i} \times U_{ab}$$

and if the condition for the maximum power $R_a = 2 \times R_i$ is applied:

$$\boldsymbol{U_{gm}} = \frac{3}{4} \times \frac{U_{ab}}{\mu}$$

The efficiency of a single ended amplifier is defined as the ratio of the useful output power to the power of the DC supply of the anode $\eta = \frac{P_{out}}{P_{ab}}$:

$$\eta = \frac{1}{2} \times \frac{1}{1+\frac{2\times R_i}{R_a}}$$

At maximum power condition: $R_a = 2 \times R_i$ → $\eta = 25\ \%$

For now, the conclusion is: the maximum power of the single ended output amplifier is $P_{out\,max} = \frac{U_{ab}^2}{16 \times R_i}$ under the condition $\mathbf{R_a = 2 \times R_i}$ **.**

In order to determine the OPTIMAL anode load (optimal load line) applicable in practice, it is necessary to perform further analyzes based on the real characteristics of the tube (the real characteristics of the tube $I_a = f(U_a)$ are highly non-linear in the region of low anode current. Tube operation in non-linear region may cause unacceptable signal distortion).

4.10.1 SINGLE ENDED TRIODE OUTPUT STAGE - LOAD LINE DESIGN

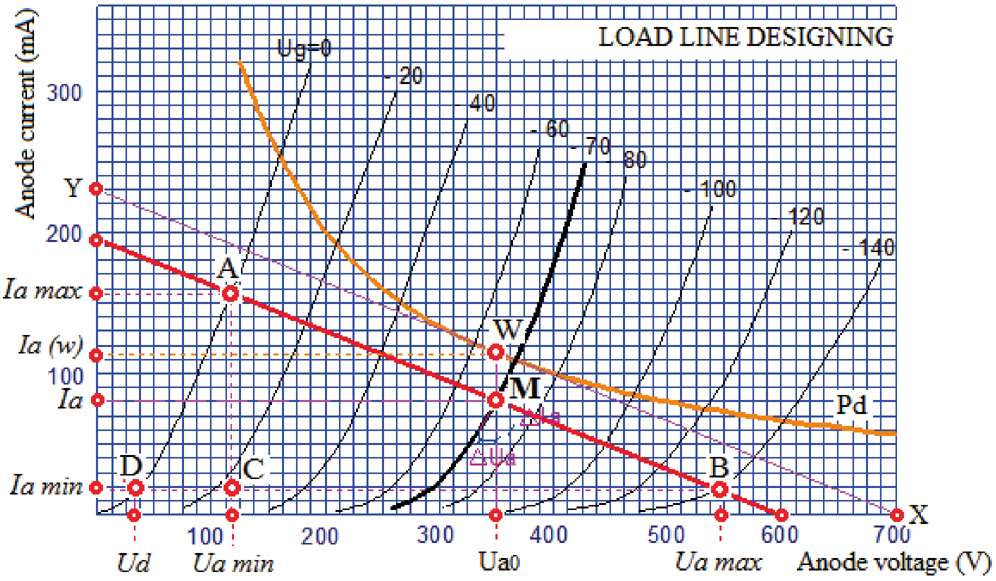


Fig. 4-45

1. Draw the power hyperbola (max anode dissipation: $\mathbf{P_d = (I_a \times U_a)}$ in the tube characteristics diagram $\mathbf{I_a = f(U_a)}$.

The power hyperbola equation is:

$$I_a = \frac{P_d}{U_a} \qquad (409.15)$$

2. For the chosen anode voltage $\mathbf{U_{a0}}$, draw the point **W** on the power hyperbola (intersection of the vertical line drawn through the point U_{a0} and the power hyperbola curve).

Coordinates of point W: $\mathbf{W\ (U_{a0}, I_{a0})}$.

3. Draw the tangent of the power hyperbola through the point W.

The equation of the tangent of the curve expressed by the function f(x) at the point (x_0, y_0) is: $f'(x_0) = \frac{y - y_0}{x - x_0}$,

applied to the diagram $I_a = f(U_a)$: $I_a'(U_{a0}) = \frac{I_a - I_{a0}}{U_a - U_{a0}}$

The first derivative of the function $I_a = \frac{P_d}{U_a}$, is: $-\frac{P_d}{U_{a0}^2} = \frac{I_a - I_{a0}}{U_a - U_{a0}}$

The equation of the tangent of the power hyperbola at the point W(U_{a0}, I_{a0}) is:

$$I_{at} = -\frac{P_d}{U_{a0}^2} \times U_a + \left(\frac{P_d}{U_{a0}} + I_{a0}\right)$$

The tangent I_{at} intersects the U_a axis at the point X: $I_{at} = 0$ and $I_{at} = 0$ and $U_{a(x)} = U_{a0} + \frac{U_{a0}^2 \times I_{a0}}{P_d}$

The tangent I_{at} intersects the I_a axis at the point Y: $U_{at} = 0$ and $U_{at} = 0$ and $I_{a(Y)} = \frac{P_d}{U_{a0}} + I_{a0}$

4. **Anode load R_a (primary impedance of the output transformer)** is determined by the **slope** of the load line:

$$\boldsymbol{R_a} = \frac{U_{a(x)}}{I_{a(y)}} = \frac{U_{a0} + \frac{U_{a0}^2 \times I_{a0}}{P_d}}{\frac{P_d}{U_{a0}} + I_{a0}} \tag{409.16}$$

5. For safe tube operation (and longer tube life), the load line and operating point (quiescent point) must be located below the power hyperbola ($0.7\ P_d \leq P_a \leq 0.95\ P_d$). For the stated reason, the real operating line of the tube (load line) is determined as a line parallel to the load line, which is determined as the tangent of the maximum power hyperbola, at the chosen anode voltage U_a and the chosen anode dissipation ($0.75\ P_d < P_a < 0.95\ P_d$). The real load R_a is equal to the anode load determined above.

6. **Output power**
   The load line intersects the **$I_a = f(U_a)$** characteristic **$U_g = 0$** at the point **$A(U_{amin}, I_{amax})$.**
   The load line intersects the **U_g** characteristic **$I_a = f(U_a)$** which is **twice as large as U_g** at the operating point **M,** at the point **$B(U_{amax}, I_{amin})$**

$$P_{out} = \frac{(U_{a\,max} - U_{a\,min}) \times (I_{a\,max} - I_{a\,min})}{8} \tag{409.17}$$

7. **Second harmonic distortion**

$$HD_2 = \frac{\frac{U_a - U_{a\,min}}{U_{a\,max} - U_a} - 1}{2 \times \left(\frac{U_a - U_{a\,min}}{U_{a\,max} - U_a} + 1\right)} \times 100\ (\%) \tag{409.18}$$

A similar result is obtained by applying the equation:

$$HD_2 = \frac{1}{2} \times \frac{I_{a\,max} + I_{a\,min} - 2 \times I_a}{I_{a\,max} - I_{a\,min}} \times 100(\%) \tag{409.19}$$

Conclusion:
The operating area of the triode used in the single ended output amplifier is limited by:

- Maximum tube dissipation hyperbola **$P_{a\,max}$ or $(0.7 \div 0.95)\ P_{a\,max}$**

- Maximum anode current $I_{a\,max}$ and the condition **$I_{a\,max} < 2 \times I_a$** (maximum I_a is less than twice the quiescent current) – condition for a class A amplifier

- Minimum I_a current that is slightly higher than the current in the extremely curved parts of the characteristics $I_a = f(U_a)$ and $U_{a\,min}$ (point at the intersection $U_g = 0V$ with $I_{a\,min}$) – the operations of the tube in the linear part of the characteristics achieves a minimum distortion of the signal.

- Maximum anode voltage determined by the minimum anode current (intersection point of the load line with characteristic Ug equal to twice the U_g at the quiescent point). The maximum anode voltage lower than twice the anode voltage at the quiescent point (**$U_{a\,max} < 2 \times U_{a0}$**).

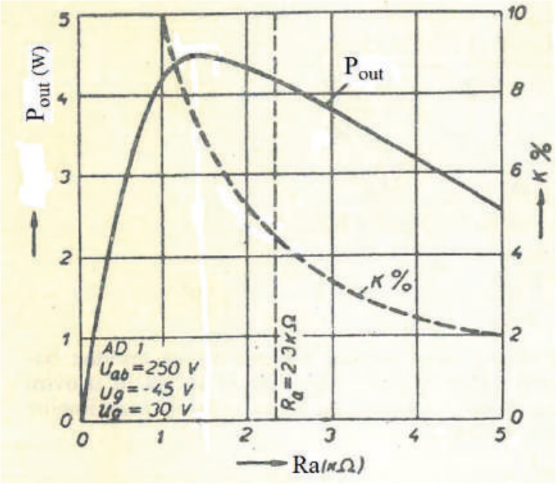


Fig. 4-46

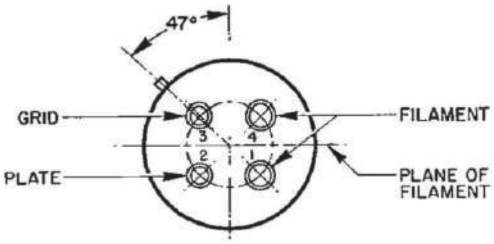
The load line is designed according to the compromise between the maximum output power (taking into account the power dissipation and the lifetime of the tube) and the acceptable signal distortion.In practice, the anode load R_a of the triode is always higher than the calculated R_a under the condition of maximum power ($R_a = 2 \times R_i$).

As it can be seen from Fig.4-46, the decrease of the harmonic distortion of the signal of a single ended triode amplifier depending on the increase of the anode load is more rapidly than depending on the decrease of the output power. In the literature, it can sometimes be found that $\mathbf{R_a = 5 \times R_i}$ is a good choice for the anode load value of the output triode of a single ended amplifier.

A triode single ended output amplifier has relatively low nonlinear distortions (both harmonic and non harmonic). Harmonic distortions of the SE amplifier are mainly caused by the second harmonics of the anode current.

4.10.2 *Examples of Triode Single Ended Output stage Load Lines*

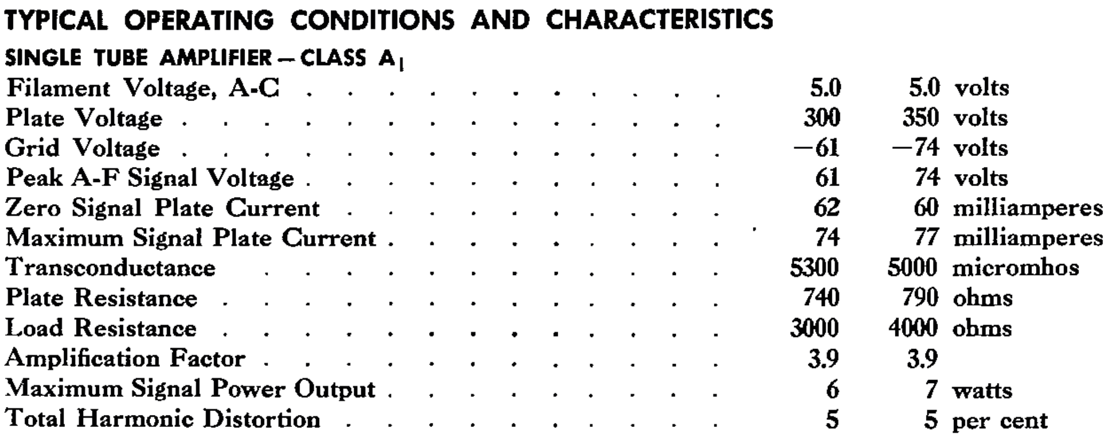
4.10.2.1 Single Ended output stage using directly heated triode 300B (Western Electric)



WE 300B Limiting Operating Conditions for Safe Operation

Plate Voltage………………………….…. 400 Volts
Plate Current ……………………..……. 100 mA
Plate Dissipation …………………..….. 36 W

Filament Voltage, AC or DC …………. 5.0 V
Filament Current ………………………. 1.2 A



TYPICAL OPERATING CONDITIONS AND CHARACTERISTICS

SINGLE TUBE AMPLIFIER — CLASS A_1

Filament Voltage, A-C	5.0	5.0	volts
Plate Voltage	300	350	volts
Grid Voltage	−61	−74	volts
Peak A-F Signal Voltage	61	74	volts
Zero Signal Plate Current	62	60	milliamperes
Maximum Signal Plate Current	74	77	milliamperes
Transconductance	5300	5000	micromhos
Plate Resistance	740	790	ohms
Load Resistance	3000	4000	ohms
Amplification Factor	3.9	3.9	
Maximum Signal Power Output	6	7	watts
Total Harmonic Distortion	5	5	per cent

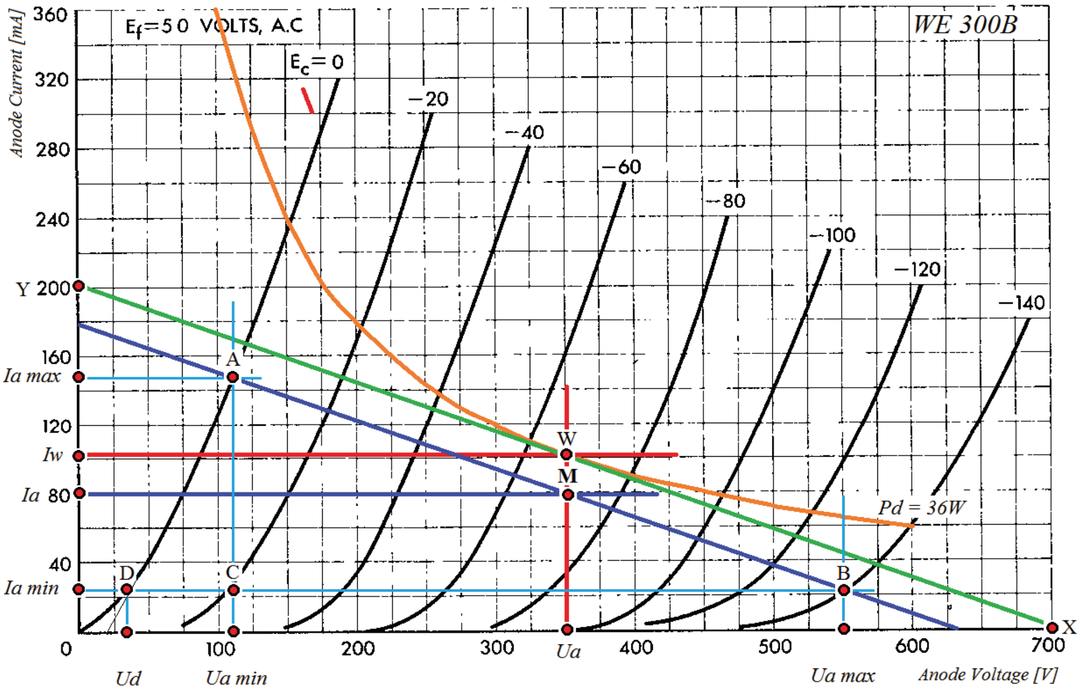


Fig. 4-47

1. Draw the power dissipation hyperbola (maximum anode power dissipation, P_d = 36 W): $I_a=\frac{P_d}{U_a}=\frac{36}{U_a}$
2. Draw the point W on the power hyperbola for the required (chosen) anode voltage (Ua = 350 V)
3. Coordinate of point W: W(350 V, 114 mA)
4. Draw a tangent line to the power hyperbola passing through point W.
5. The tangent line intersects the U_a axis at the point X (698, 0) i.e. 698 V (the ***voltage at the point where the current is equal to 0 must fulfill the condition: it must be less than twice the anode operating voltage***, i.e. 698 V < 2 x 350 V).
6. The tangent line intersects the I_a axis at point Y (0 V, 200 mA), i.e. 200 mA.
7. Calculation of R_a (Load – impedance of the transformer primary): R_a is equal to the slope of the load line: $\boldsymbol{R_a}$ = 698V / 200mA = ***3490*** Ω
8. Ia at point W, (103mA) is out of the limit for safe operation (100mA) of the tube. Also, for reasons of safety and the lifetime of the tube it is recommended that the tube operate at 0.7 to 0.8 of the maximum anode power dissipation. For example: P = 28W (that is 77 % of $P_{d\,max}$ = 36W, i.e. I_a = 28 W / 350 V = 80 mA). So, the new working point (quiescent point) M has the coordinate: ***M (350 V, 80 mA***).
9. Draw a straight line parallel to the X - Y line through point M (350 V, 80 mA).
10. The new line intersects the Ua axis at approximately 605 V.
11. The new point M is located on the characteristic $\boldsymbol{U_g = -70}$ ***V***.

12. *The new working line (load line) intersects the characteristic $U_g = 0$ at point A (110 V, 148 mA), and the symmetrical characteristic $U_g = -140$ V at point B (550 V, 20 mA):*
    $U_{amax} = 550$ V, $U_{amin} = 110$ V, $I_{amax} = 148$ mA, $I_{amin} = 20$ mA.
13. *The straight line $I_a = 20$ mA intersects the characteristic $U_g = 0$ V at point D (30 V, 20 mA)*
14. *The new working line (load line) has the same slope as the first constructed load line (7.) i.e. $R_a = 3490\ \Omega$.*
15. *The internal resistance of the tube at the working point (quiescent point) M (350 V, 80 mA) is approximately:*
    $R_i = \Delta U_a / \Delta I_a \mid [U_g = -70\text{ V}] = 14.5\text{ V} / 20\text{ mA} = 725\ \Omega$.
16. *Output power:*

$$P_{out} = \frac{1}{2} \times \frac{(U_a - U_d)^2}{(R_a + 2 \times R_i)^2} \times R_a = \frac{1}{2} \times \frac{(350 - 30)^2}{(3490 + 2 \times 725)^2} \times 3490 = 7.32\ W$$

Also, the area of the triangle ABC is proportional to the output power:

$$P_{out} = \frac{(U_{a\,max} - U_{a\,min}) \times (I_{a\,max} - I_{a\,min})}{8} = \frac{(550 - 110) \times (0.148 - 0.02)}{8} = 7.04\ W$$

17. *The necessary amplitude of the driving voltage applied to the control grid (g1)to produce the nominal output power:*

$$U_{g\,m} = \frac{1}{\mu} \times \frac{R_i + R_a}{R_i + 2 \times R_i} \times (U_a - U_{a\,min}) = \frac{1}{3.9} \times \frac{725 + 3490}{3490 + 2 \times 725} \times (350 - 30) = 70\ V$$

18. *Efficiency:*

$$\eta = \frac{1}{2} \times \frac{1}{1 + \frac{2 \times R_i}{R_a}} = \frac{1}{2} \times \frac{1}{1 + \frac{2 \times 725}{3490}} = 0.35\ or\ 35\ \%$$

19. *An approximately equal value of the output power is obtained by applying the equation:*

$$P_{out} = \frac{1}{2} \times \left(\frac{\mu \times U_{g\,m}}{R_i + R_a}\right)^2 \times R_a = \frac{1}{2} \times \left(\frac{3.9 \times 70}{725 + 3490}\right)^2 \times 3490 = 7.32\ W$$

20. *Second harmonic distortion:*

$$HD_2 = \frac{\frac{U_a - U_{a\,min}}{U_{a\,max} - U_a} - 1}{2 \times \left(\frac{U_a - U_{a\,min}}{U_{a\,max} - U_a} + 1\right)} \times 100\ (\%) = \frac{\frac{350-110}{550-350} - 1}{2 \times \left(\frac{350-110}{550-350} + 1\right)} \times 100 = 4.5\ \%$$

An approximately equal value of the second harmonic distortion is obtained by applying the equation:

$$HD_2 = \frac{1}{2} \times \frac{I_{a\,max} + I_{a\,min} - 2 \times I_a}{I_{a\,max} - I_{a\,min}} \times 100\ (\%) = \frac{1}{2} \times \frac{0.148 + 0.02 - 2 \times 0.08}{0.148 - 0.02} = 3.125\ \%$$

The electric diagram (schematic) of the output stage calculated above is:

Working point (quiescent point) is: M (350 V, 80 mA), - U_{gk} = 70 V.

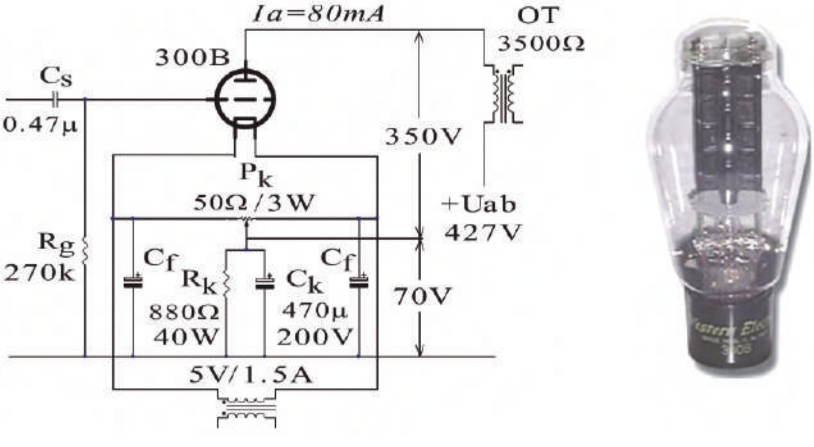


Fig. 4-48

*300B is directly heated cathode triode, i.e. the filament itself is the cathode and the filament is heated using a separate power supply source (according to the tube data sheet): 5V_{AC}/1.5A - separate transformer or separate winding of the transformer used in the amplifier power supply circuit). (*DC 5V_{DC}/1.5A voltage source can also be used).*

Grid bias is provided by an automatic grid bias circuit, i.e., by using a resistor in the cathode circuit.
Cathode resistor:

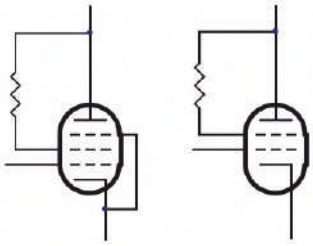
$$R_k = \frac{U_{gk}}{I_a} = \frac{70V}{0.08A} = 875\,\Omega \approx 880\,\Omega$$

Capacitance C_k is calculated using the equation: $2 \times \pi \times f_{Low} \times C_k \times R_k > 30 \rightarrow C_k > 279\,\mu F$ (f_{Low} = 20 Hz)

DC high voltage power supply:

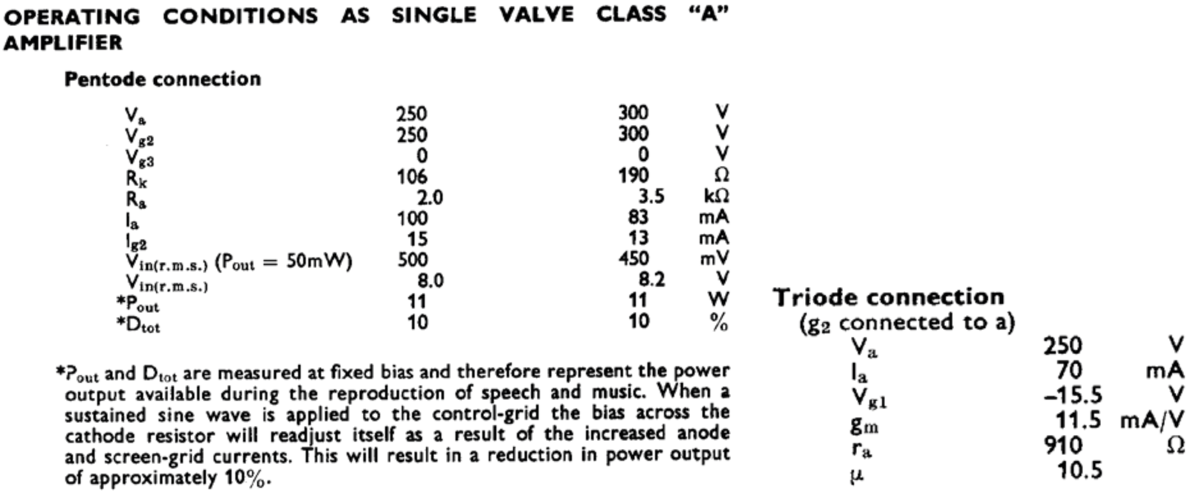
$$U_{ab} = U_k + U_{ak} + U_{OT-D} = 70\,V + 350\,V + R_{OT-D} \times I_a \approx 427\,V$$

Single-Ended Triode output amplifiers can be built using classic power pentodes or a power beam tetrodes connected as triodes (triode mode - screen grid connected to the anode).



4.10.2.2 Single-Ended output stage: EL34 power pentode connected as triode–triode mode

Factory Data:
EL34 Limiting Operating Conditions for Safe Operation
(Triode connection)
Plate Voltage... 600 Volts
P_a + g2 max. (V_a = 500V) .... 30 W
Heater: 6.3V / 1.5A



OPERATING CONDITIONS AS SINGLE VALVE CLASS "A" AMPLIFIER

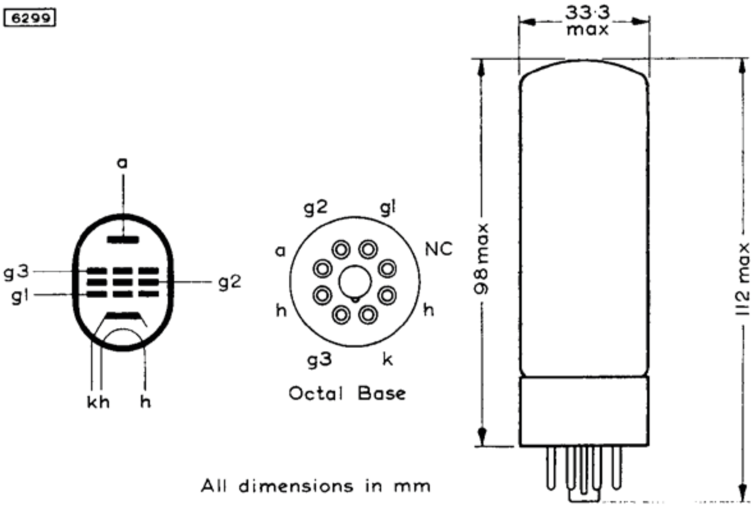
Pentode connection

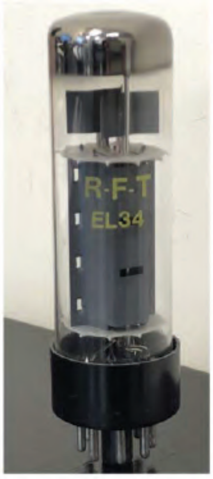
V_a	250	300	V
V_{g2}	250	300	V
V_{g3}	0	0	V
R_k	106	190	Ω
R_a	2.0	3.5	kΩ
I_a	100	83	mA
I_{g2}	15	13	mA
$V_{in(r.m.s.)}$ (P_{out} = 50mW)	500	450	mV
$V_{in(r.m.s.)}$	8.0	8.2	V
*P_{out}	11	11	W
*D_{tot}	10	10	%

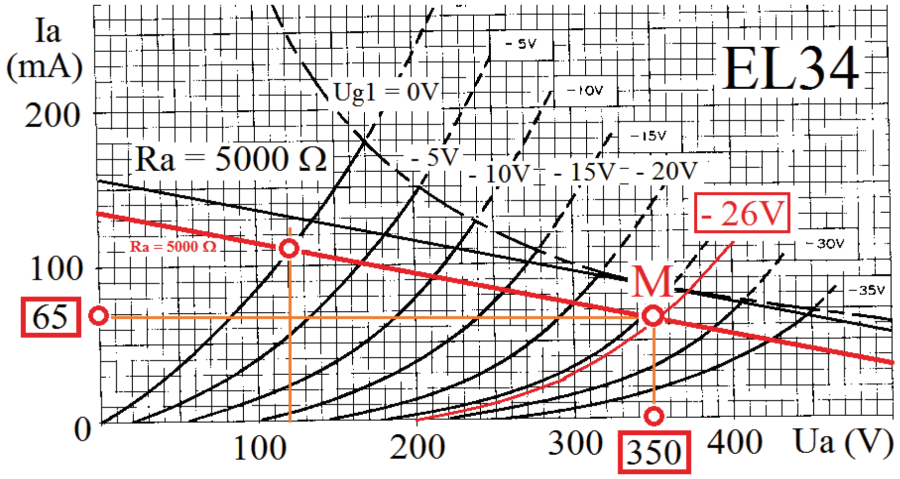
*P_{out} and D_{tot} are measured at fixed bias and therefore represent the power output available during the reproduction of speech and music. When a sustained sine wave is applied to the control-grid the bias across the cathode resistor will readjust itself as a result of the increased anode and screen-grid currents. This will result in a reduction in power output of approximately 10%.

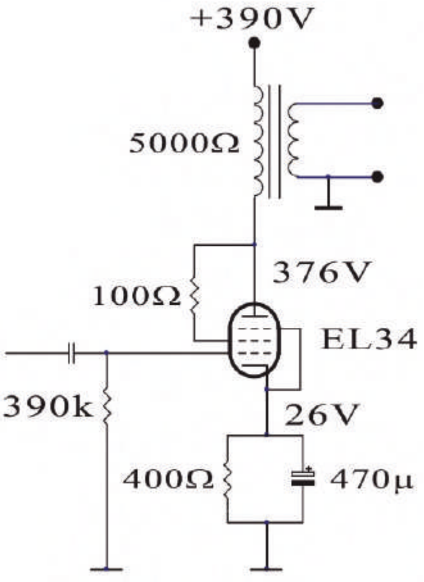
Triode connection
(g_2 connected to a)

V_a	250	V
I_a	70	mA
V_{g1}	–15.5	V
g_m	11.5	mA/V
r_a	910	Ω
μ	10.5	









Quiescent point M (350 V, 65 mA)

U_{gk} = - 26 V

R_k = 400 Ω

R_a = 5000 Ω

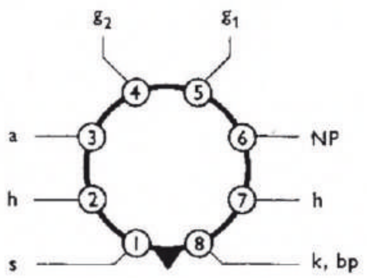
U_{ab} = 390 V

P_{out} = 5.3 W

Fig. 4-49

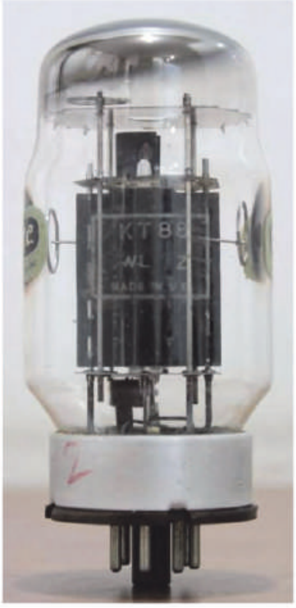
4.10.2. 3 Single-Ended output stage: KT88 power beam tetrode connected as triode–triode mode

BASE CONNECTIONS AND VALVE DIMENSIONS



Base: Metal shell wafer octal
Bulb: Tubular

Max. overall length: 125mm
Max. seated length: 110mm
Max. diameter: 52mm



HEATER

V_h	6·3	V
I_h	1·6 (approx)	A

MAXIMUM RATINGS

	Absolute	Design Maximum	
V_a	800	800	V
V_{g2}	600	600	V
$V_{a,g2}$	600	600	V
$-V_{g1}$	200	200	V
p_a	42	35	W
p_{g2}	8	6	W
p_{a+g2}	46	40	W
I_k	230	230	mA
V_{h-k}	250	200	V
T_{bulb}	250	250	°C
R_{g1-k} (cathode bias)			
$p_{a+g2} \leq 35W$	470		kΩ
$p_{a+g2} > 35W$	270		kΩ
R_{g1-k} (fixed bias)			
$p_{a+g2} \leq 35W$	220		kΩ
$p_{a+g2} > 35W$	100		kΩ

KT88 Limiting Operating Conditions for Safe Operation (according to the data sheet):

- Maximum anode, screen grid voltage..........................600 V
- Maximum anode dissipation..42 W
- Maximum anode current (for self biasing)..................230 mA
- Amplification factor (μ) (triode connection)…….........8

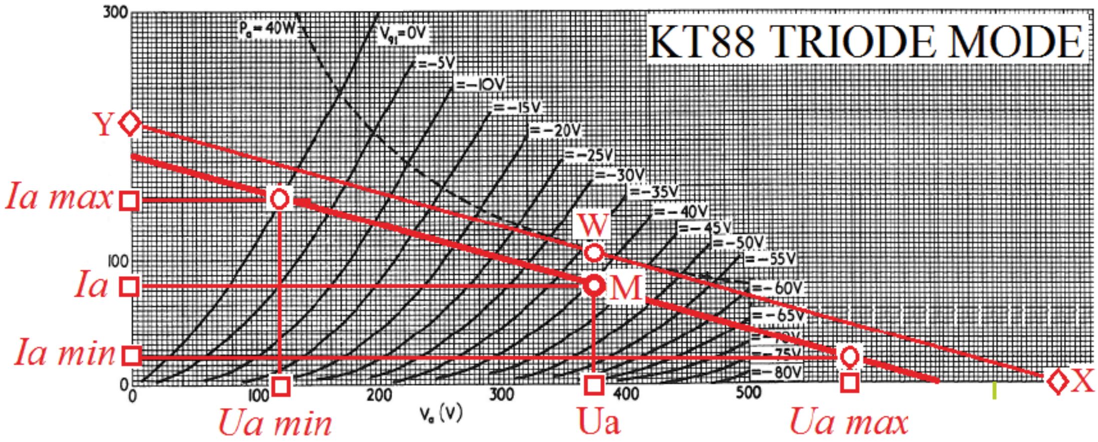


Fig. 4-50

Set of characteristics $I_a = f(U_a)$, U_g = const.(anode characteristics) is shown in Fig.4-50.

1. *Draw the power dissipation hyperbola (maximum anode power dissipation, P_d = 42W):* $I_a = \frac{P_d}{U_a} = \frac{42}{U_a}$
2. *Draw the point W on the power hyperbola for the required (chosen) anode voltage (375V)*
3. *Coordinate of point W: W(375 V, 115 mA)*
4. *Draw a tangent line to the power hyperbola passing through point W.*
5. *The tangent line intersects the Ua axis at the point X (760, 0) i.e. 760 V*
6. *The tangent line intersects the Ia axis at the point Y (0, 215), i.e. 215 mA.*
7. *Calculation of R_a (Load – impedance of the transformer primary): R_a is equal to the slope of the load line: R_a = 760 V / 215 mA = 3535 Ω.*
8. *Set (draw) the new working (quiescent) point M on the characteristic $U_g \approx -42$ V. M (375 V, 76 mA).*
9. *Draw a straight line parallel to the X-Y line through point M (375 V, 76 mA).*
10. *The new working (load) line intersects the characteristic U_g = 0V at point A (122V, 150mA), and the symmetrical characteristic U_g = - 84 V at point B (590 V, 18 mA) approximately:*
    U_{amax} = 590 V, U_{amin} = 122 V, I_{amax} = 150 mA, I_{amin} = 18 mA.
11. *The straight line I_a =18 mA intersects the characteristic U_g = 0 V at point D (25 V, 18 mA)*
12. *The new working line (load line) has the same slope as the first constructed load line i.e. R_a = 3535 Ω.*
13. *The internal resistance of the tube at the working point (quiescent point) M (375 V, 76 mA) is approximately: $R_i = \Delta U_a / \Delta I_a \mid [U_g = -42\ V]$ = 930 Ω.*
14. *Output power:*

$$P_{out} = \frac{1}{2} \times \frac{(U_a - U_d)^2}{(R_a + 2 \times R_i)^2} \times R_a = \frac{1}{2} \times \frac{(375-25)^2}{(3500+2\times930)^2} \times 3500 = 7.46\ W$$

Also, the area of the triangle ABC is proportional to the output power:

$$P_{out} = \frac{(U_{a\,max} - U_{a\,min}) \times (I_{a\,max} - I_{a\,min})}{8} = \frac{(590-122) \times (0.15 - 0.018)}{8} = 7.72\ W$$

15. *The necessary amplitude of the driving voltage applied to the control grid (g1)to produce the nominal output power:*

$$U_{gm} = \frac{1}{\mu} \times \frac{R_i + R_a}{R_a + 2 \times R_i} \times (U_a - U_d) = \frac{1}{8} \times \frac{630 + 3500}{3500 + 2 \times 930} \times (375 - 25) = 36\ V$$

16. *Efficiency:*

$$\eta = \frac{1}{2} \times \frac{1}{1 + \frac{2 \times R_i}{R_a}} = \frac{1}{2} \times \frac{1}{1 + \frac{2\times930}{3500}} = 0.326 = 32.6\%$$

17. *An approximately equal value of the output power is obtained by applying the equation:*

$$P_{out} = \frac{1}{2} \times \left(\frac{\mu \times U_{gm}}{R_i + R_a}\right)^2 \times R_a = \frac{1}{2} \times \left(\frac{8 \times 36}{930 + 3500}\right)^2 \times 3500 = 7.34\ W$$

18. *Second-harmonic distortion:*

$$HD_2 = \frac{\frac{U_a - U_{a\,min}}{U_{a\,max} - U_a} - 1}{2 \times \left(\frac{U_a - U_{a\,min}}{U_{a\,max} - U_a} + 1\right)} \times 100(\%) = \frac{\frac{375-122}{590-375} - 1}{2 \times \left(\frac{375-122}{590-375} + 1\right)} \times 100(\%) = 0.04 \times 100(\%) = 4\ \%$$

An approximately equal value of the second harmonic distortion is obtained by applying the equation:

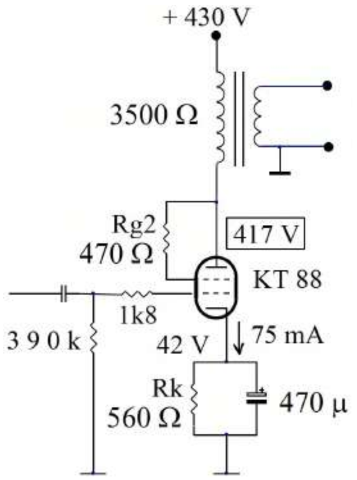
$$HD_2 = \frac{1}{2} \times \frac{I_{amax} + I_{a\,min} - 2 \times I_a}{I_{a\,max} - I_{a\,min}} \times 100(\%) = \frac{1}{2} \times \frac{150+18-2\times76}{150-18} \times\times 100(\%) = 6\ \%$$

$R_k = U_{gk}$ / I_a = 42 V / 0.076 mA = 552.6 Ω ≈ 560 Ω

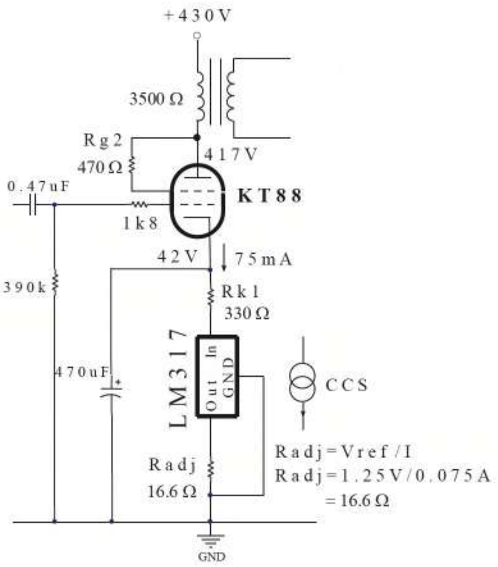
$P_{Rk} = R_k \times I^2_a$ = 560 Ω x (0.076 mA)^2 = 3.24 W

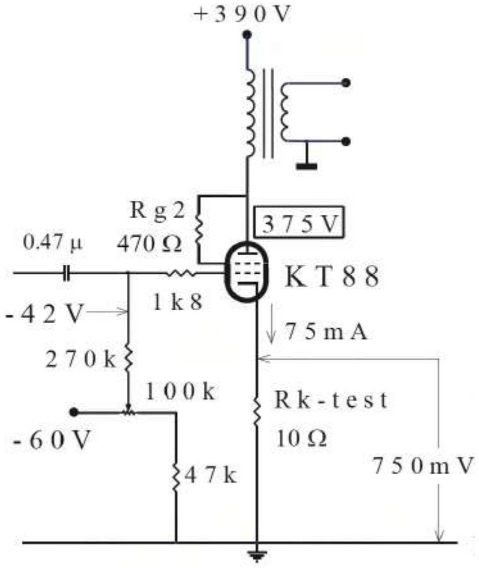
*R_k = 560 Ω / 12 W

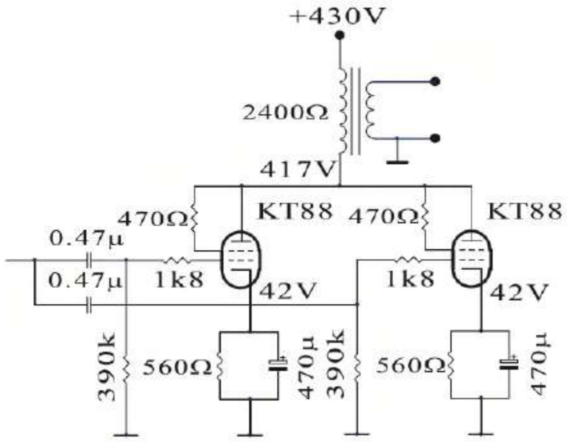
$U_{ab} = U_k + U_{ak} + I_a \times R_{DC-OT}$ = 42 V + 375 V + 0.075 mA x R_{DC-OT} ≈ 430 V



The bias circuit of the output tube of the of the Single-Ended Triode output amplifier can be built using other design solutions of the bias circuit (fixed bias, constant current source), and not only using the automatic bias circuit applied in the above examples.







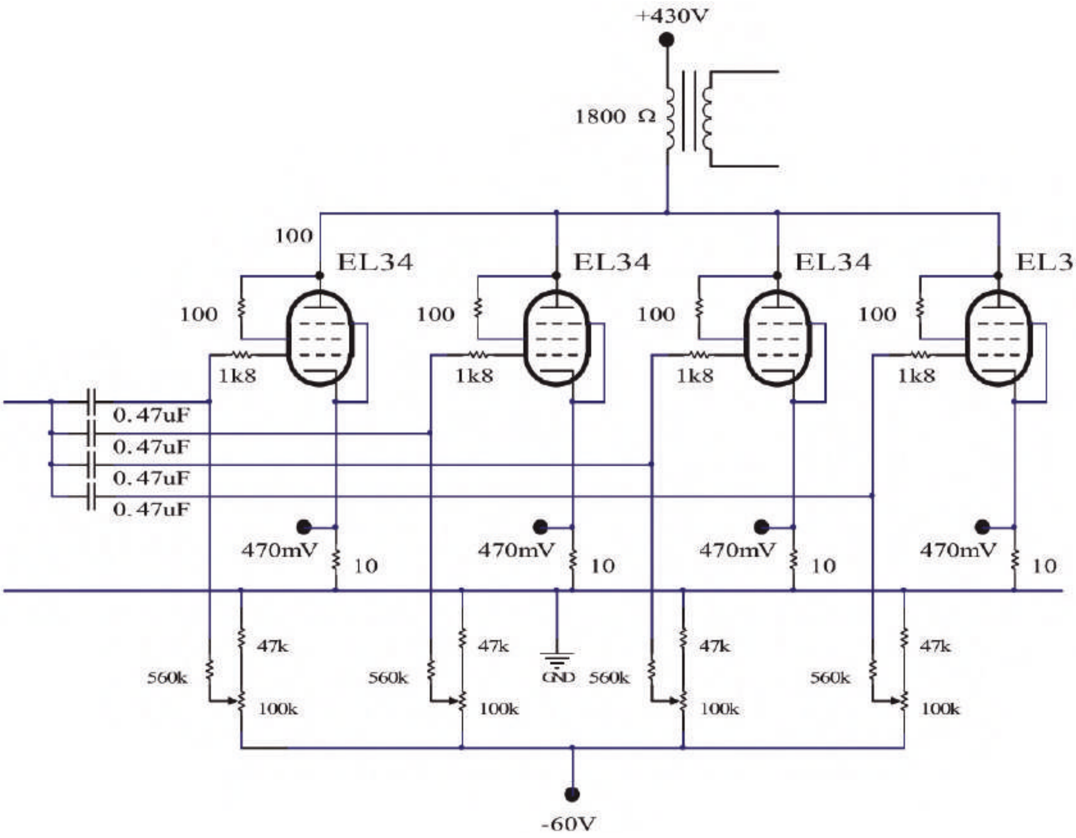
Single-Ended Triode output amplifiers of higher output powers can be designed using the parallel connection of two or more triodes (PSET): **# It is very hard to find several tubes with identical characteristics. So, if two or more output tubes are connected in parallel, it is necessary to use independent bias circuits for each tube.**

The effective anode resistance (internal tube resistance) of the PSET is: R_i (one-tube triode SE amplifier) / number of tubes connected in parallel.

The load resistance of the PSET is: R_a (one-tube triode SE amplifier) / number of tubes connected in parallel.

The total anode current of the PSET is: I_a (one-tube triode SE amplifier) x number of tubes connected in parallel.

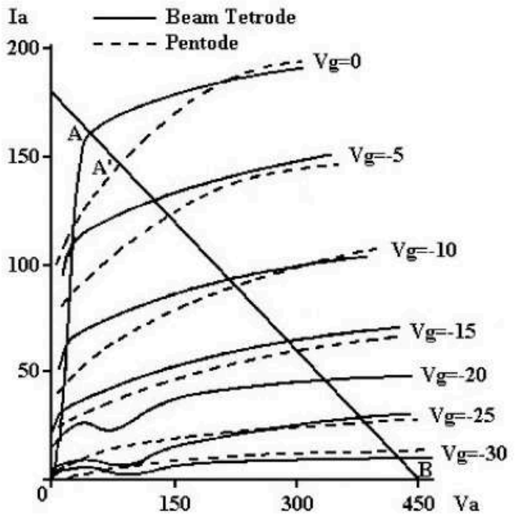
The total harmonic distortion of the PSET is approximately equal to the harmonic distortion of a single-tube SE triode amplifier. The Input (driving) voltage of the PSET is equal to the driving voltage of a single- tube SE triode amplifier.

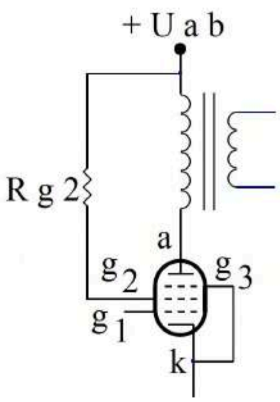


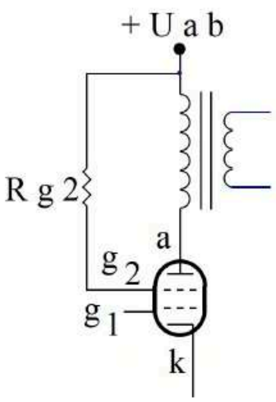
4.10.3 SINGLE-ENDED PENTODE AND TETRODE OUTPUT STAGE

SE Power Pentode and Power Beam Tetrode output amplifiers produce higher output power than SE triode amplifiers for equal output tube anode power dissipation, i.e. the power efficiency of SE Power Pentode and Power Beam Tetrode output amplifiers is higher than the power efficiency of SE triode output amplifiers. The sensitivity is also higher (at equal output powers the driver signal (signal applied to the control grid) of the SE pentode and beam tetrode amplifier is several times lower than the driver signal of the triode output amplifier). The non-linear distortion of the SE Power Pentode and Power Beam Tetrode output amplifier is higher than the non-linear distortion of the SE triode amplifier.

The anode resistance (internal resistance) of the pentode and beam tetrode is very high (several tens of kΩ).







Graphical Analysis - Single Ended Pentode Output Stage Load line

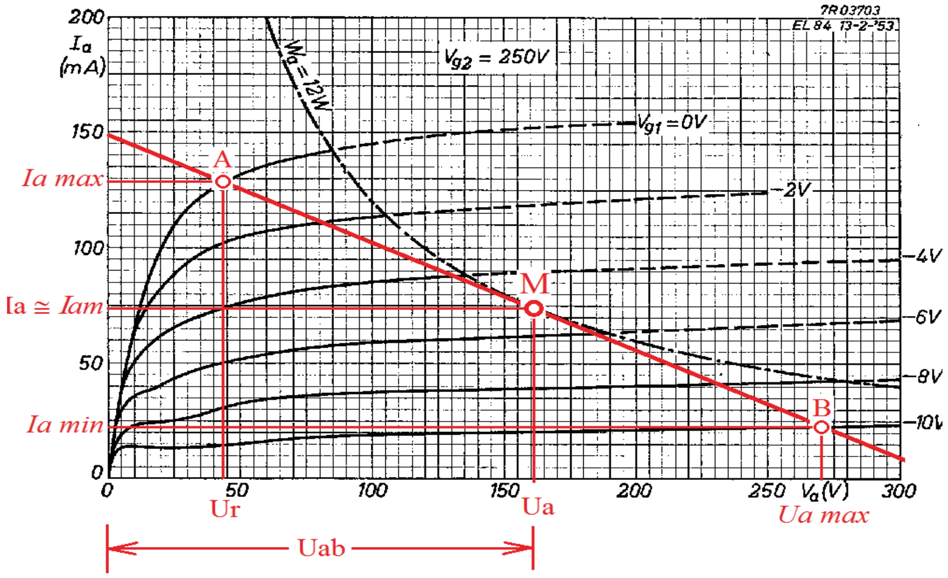


Fig. 4-51

The maximum output power is achieved under the condition: the intersection of the load line and the characteristic Ia = f(Ua)│Ug = 0 is located at the knee of the curve Ug = 0 (the part of the characteristic Ug = 0 that is maximally curved).

Note:
Fig. 4-51: Optimal load line AMB (minimum harmonic distortion of the output signal – no signal distortion): AM equal to MB. [M (U_a, I_a) is the quiescent point – the operating point].

Optimal anode load: $R_{Load} = \frac{U_a}{I_a}$ (in practice: $R_{Load} = 0.9 \times \frac{U_a}{I_a}$)

Output power: $P_{out} = \frac{1}{2} \times U_a \times I_a = \frac{1}{2} \times (U_{ab} - U_r) \times I_a$

Efficiency: $\eta = \frac{P_{out}}{P_{ab}} = \frac{U_a \times I_a}{U_{ab} \times I_a} = \frac{1}{2} \times \left(1 - \frac{U_r}{U_{ab}}\right)$ (in practice: η = (35 ÷ 45) %)

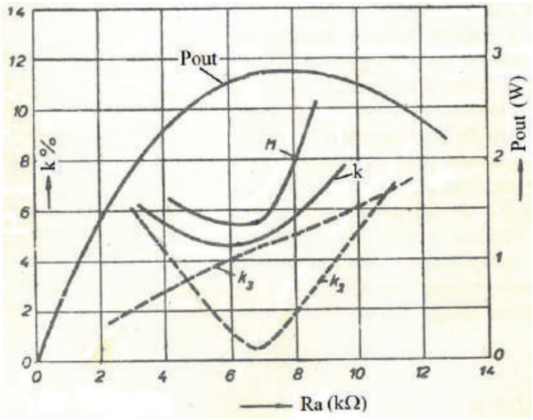


Fig. 4-52

It is very important to match the conditions for achieving maximum output power depending on R_{Load} and the conditions for achieving minimum non-linear distortion of the output signal. Pentodes distort the amplified signal more than triodes. The second and third harmonics of the output signal are high which means higher non-harmonic non-linear distortion than those of SE triode amplifier. Fig. 4-52 is a diagrams of output power P_{out} , second (k_2) and third (k_3) harmonic distortion, total distortion (k) and non harmonic distortion (M) as a function of R_{Load} . Analyzing the digram

Fig. 4-52 it can be observed that the non-linear distortion is minimal for the value of R_{Load} coresponding to the minimal distortion (almoust zero) of the second harmonic.

The amplitude of the second harmonic of the anode current is:

$$I_{2\omega m}=\frac{I_{amax}+I_{amin}-2\times I_a}{4}$$

Also: $I_{amax}\approx 2\times I_a - I_{amin}$ so it can be concluded that the non-linear distortion is minimal for the specified position of the load line.

 # Total Harmonic Distortion (THD) of a SE pentode power amplifier can be reduced by applying negative feedback (NFB).

Power supply of the screen grid (g2):

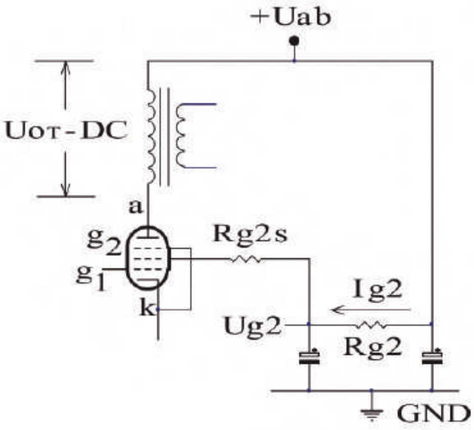
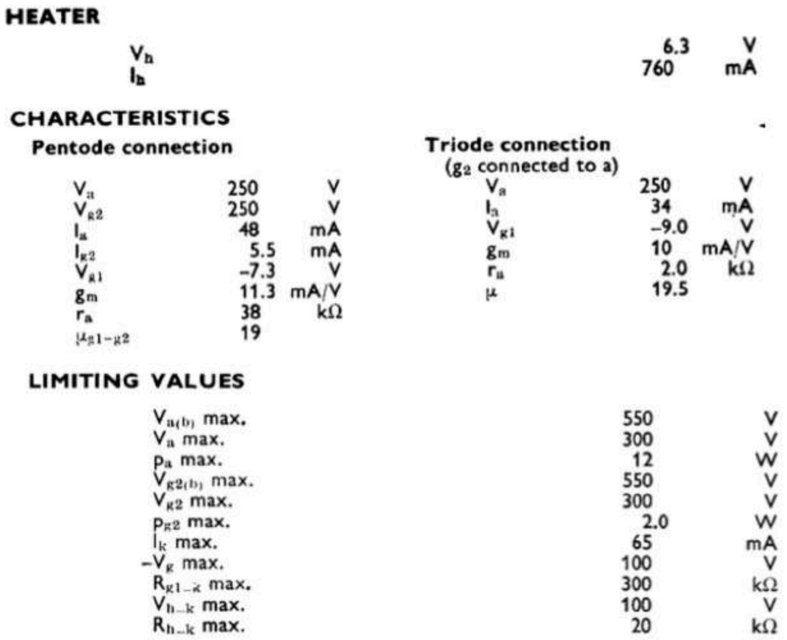


Fig. 4-53

It is not necessary to use a separate power supply unit to power the screen grid. The anode power supply of the amplifier can be used to power the screen grid, which is common in practice (Fig.4-53). It is necessary to consult the tube manufacturer data sheet to determine the screen grid current I_{g2} (for the determined anode voltage, or control grid voltage).

It is known that if the output pentode of the amplifier is overdriven (the pentode operates in the area of low anode voltages), the screen grid current rises rapidly and the tube can be damaged. To overcome this problem, a screen grid stopper resistor $\mathbf{R_{g2S}}$ is connected in the screen grid circuit, acting to reduce the screen grid voltage as the screen grid current rises, thereby limiting the power dissipation of the screen grid. In practice, the resistance of R_{g2S} is between 100 Ω and 1000 Ω for most power pentodes used in SE amplifiers.

4.10.3.1 Single-Ended output stage: EL84 (6BQ5) medium power pentode



HEATER

V_h	6.3	V
I_h	760	mA

CHARACTERISTICS

Pentode connection

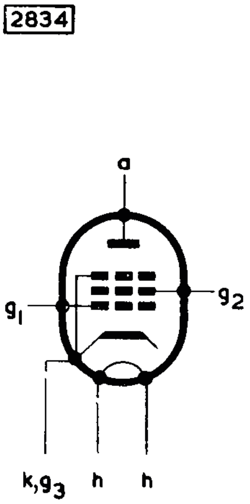
V_a	250	V
V_{g2}	250	V
I_a	48	mA
I_{g2}	5.5	mA
V_{g1}	-7.3	V
g_m	11.3	mA/V
r_a	38	kΩ
μ_{g1-g2}	19	

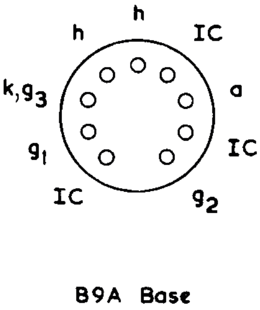
Triode connection
(g₂ connected to a)

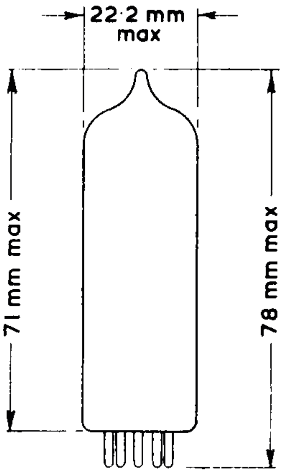
V_a	250	V
I_a	34	mA
V_{g1}	-9.0	V
g_m	10	mA/V
r_a	2.0	kΩ
μ	19.5	

LIMITING VALUES

$V_{a(b)}$ max.	550	V
V_a max.	300	V
p_a max.	12	W
$V_{g2(b)}$ max.	550	V
V_{g2} max.	300	V
p_{g2} max.	2.0	W
I_k max.	65	mA
$-V_g$ max.	100	V
R_{g1-k} max.	300	kΩ
V_{h-k} max.	100	V
R_{h-k} max.	20	kΩ







Anode voltage (chosen or required): $U_a = 270\ V$

Anode power dissipation (chosen or required): $P_d \le 0.8 \times P_{d\,max}$, $P_d \le 0.8 \times 12W$, $P_d \le 9.6\ W$

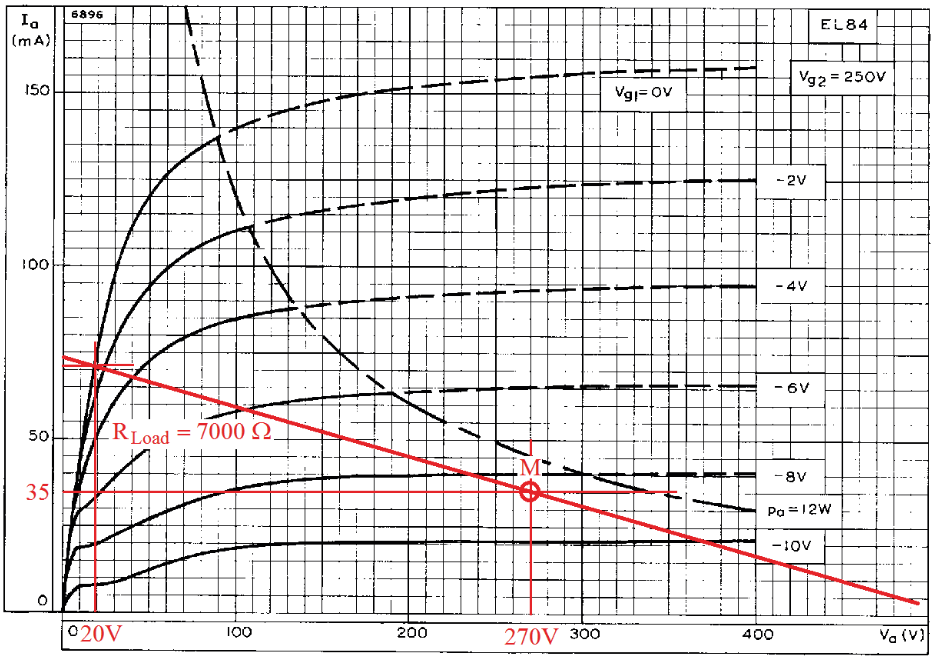
Anode quiescent current: $I_a = P_d / U_a = 9.6\ W / 270\ V = 35.5\ mA \approx 35\ mA$

Quiescent point M (270 V, 35 mA)

Using the diagram of anode characteristics $I_a = f(U_a) \rightarrow U_{gk} = -8.5\ V$

Bias voltage of the control grid: $U_{gk} = -8.5\ V$

Screen grid current: $I_{g2} \approx 4.5\ mA$



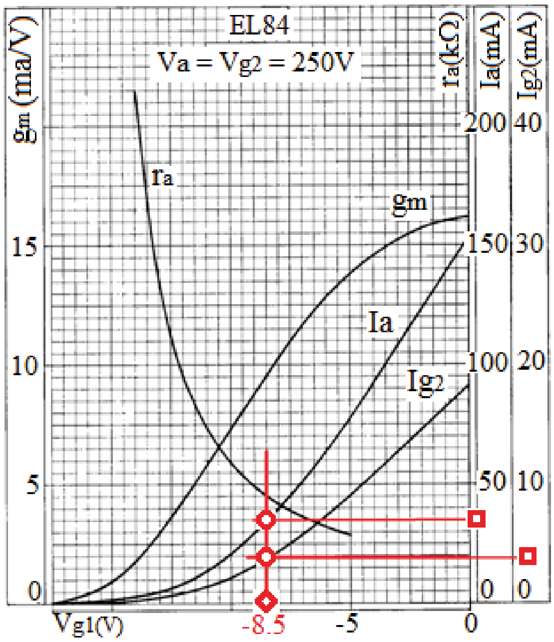
Calculation of the Anode Load: $R_{Load}=0.9\times\frac{U_a}{I_a}=0.9\times\frac{270}{0.035}=6943\,\Omega\approx 7000\,\Omega$

Draw Anode Load line R_{Load} ***= 7000 Ω through the quiescent point M (270 V, 35 mA)***

Calculate $R_k = U_{gk} / (I_a + I_{g2}) = 8.5\ V / (35\ mA + 4\ mA) =$ ***218 Ω ≈ 22 Ω***

EL84 (SE PENTODE)

Screen grid (g_2 .) Circuit design and calculation:



Tube manufacturers also publish screen grid current data: ***screen grid current as a function of control grid voltage***

Using such of diagram for EL84 and $\boldsymbol{I_{g2}\ \approx\ 4\ mA}$
($U_a = U_{g2} = 250\ V$):
The anode circuit power supply voltage is:

$$U_{ab} = U_k + U_{ak} + U_{OT\text{-}DC}$$

$$U_{ab} = 8.5 + 270 + 6.5 = 285V$$

** $U_{OT\text{-}DC}$ is approximately 6.5 V*
It is necessary to adjust the screen grid power supply voltage to a value of ***258.5 V*** *($U_k + U_{g2} = 8.5\ V + 250\ V$):*

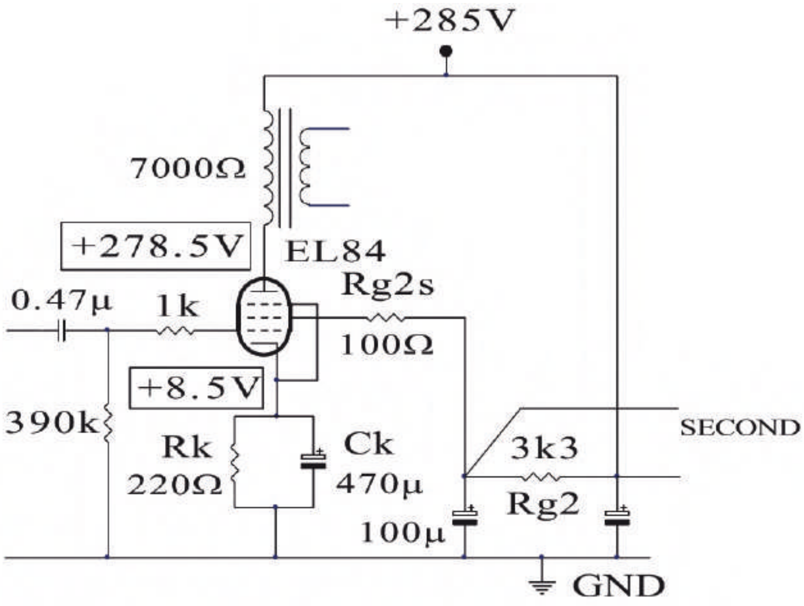
$$\boldsymbol{R_{g2\ total}} = R_{g2} = \frac{U_{ab} - 258.5\,V}{I_{g2}} = \frac{285 - 258.5}{4\,mA} = 6625\,\Omega$$

If a common power supply circuit is used for both channels of the amplifier: $\boldsymbol{R_{g2} = 3.3\ k\Omega}$ ***(two separate screen grid stoppers*** $\boldsymbol{R_{g2S1} = R_{g2S2} = 100\ \Omega}$ ***are also used).***

Calculation of output power

$$P_{out} = \frac{1}{2} \times (U_{ab} - U_r) \times I_a = \frac{1}{2} \times (270\,V - 20\,V) \times 0.035\,A = 4.375\,W$$

Or $$P_{out} = \frac{1}{2} \times \frac{(U_a - U_r)^2}{R_{Load}} = \frac{1}{2} \times \frac{(270\,V - 20\,V)^2}{7000\,\Omega} = 4.46\,W$$



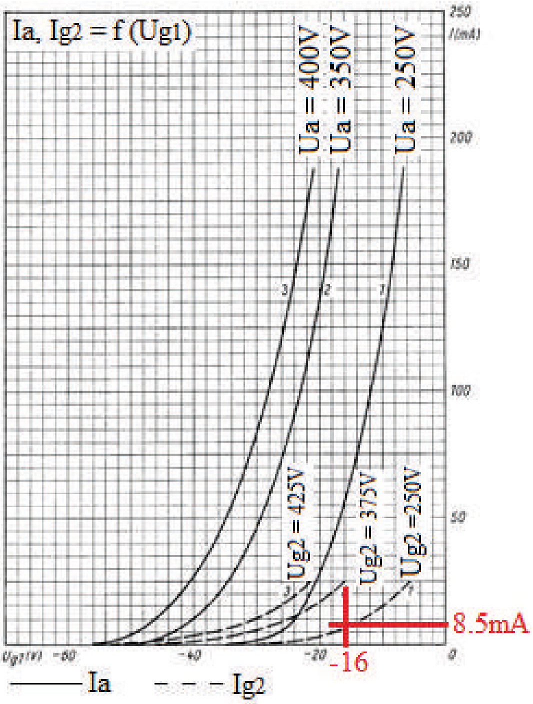
4.10.3.2 Single-Ended output stage: EL34 power pentode

Anode voltage (chosen or required): $U_a = 320\ V$

Anode power dissipation (chosen or required): $P_d \le 0.8 \times P_{d\,max}$, $P_d \le 0.8 \times 25\ W$, $P_d \le 20\ W$

Anode quiescent current: $I_a = P_d / U_a = 20\ W / 320\ V = 62.5\ mA \approx 60\ mA$

Quiescent point M (320 V, 60 mA)



Using the diagram $I_a = f(U_a) \rightarrow U_{gk} = -16\ V$,

Bias voltage: $U_{gk} = -16\ V$

Screen grid current: $I_{g2} \approx 8.5\ mA$

Calculation of anode load:

$$R_{Load} = 0.9 \times \frac{U_a}{I_a} = 0.9 \times \frac{320}{0.06} = 4800\,\Omega$$

Draw Anode Load line R_{Load} = 4800 Ω through the quiescent point M (320 V, 60 mA)

Calculation of $R_k = U_{gk} / (I_a + I_{g2})$
= 16 V / (60 mA + 8.5 mA) = **233.5 Ω ≈ 240 Ω**

The anode circuit power supply voltage is:

$$U_{ab} = U_k + U_{ak} + U_{OT-DC}$$

$$U_{ab} = 16 + 320 + 14 = 350\,V$$

*U_{OT-DC} is approximately 14V

It is necessary to adjust the screen grid power supply voltage to a value of ***266 V****.*

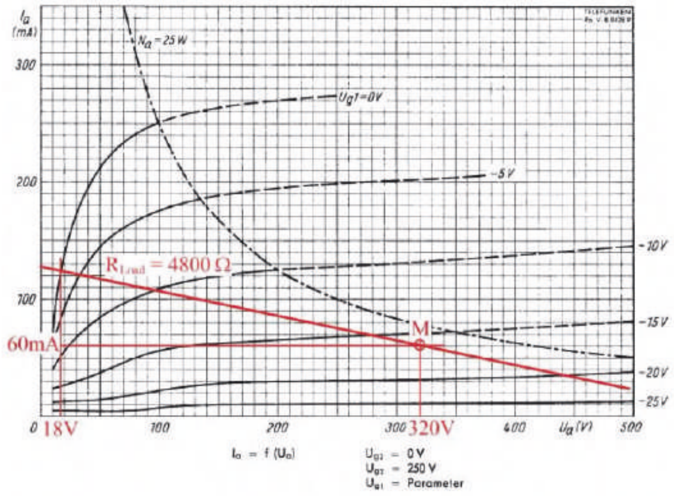
$$R_{g2total} = R_{g2} = \frac{U_{ab} - 266}{I_{g2}} = \frac{350\,V - 266\,V}{8.5\,mA} = 9882\,\Omega$$

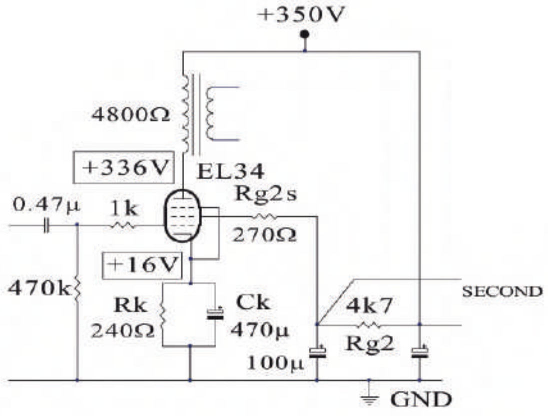
If a common power supply circuit is used for both channels of the amplifier: R_{g2} = 4.7 k Ω (two separate screen grid stoppers R_{g2S1} = R_{g2S2} = 270 Ω are also used).

Calculation of output power:

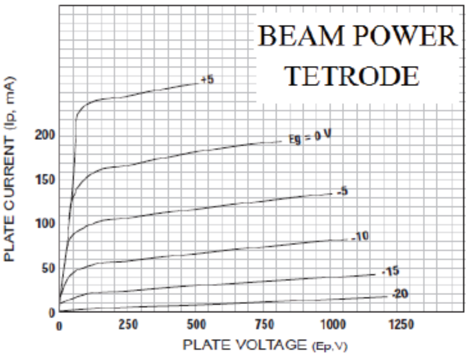
$$P_{out} = \frac{1}{2} \times (U_{ab} - U_r) \times I_a = \frac{1}{2} \times (320\,V - 18\,V) \times 0.06\,A = 9\,W$$

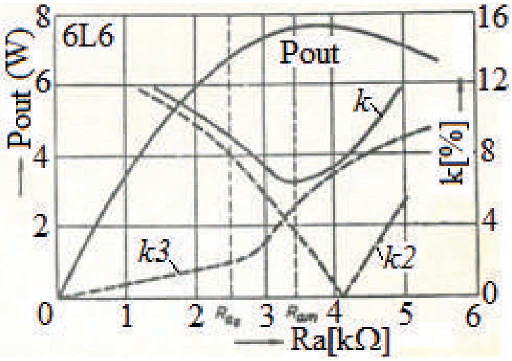
or, $$P_{out} = \frac{1}{2} \times \frac{(U_a - U_r)^2}{R_{Load}} = \frac{1}{2} \times \frac{(320\,V - 18\,V)^2}{4800\,\Omega} = 9.5\,W$$





4.10.3.3 Single Ended power beam tetrode output stage

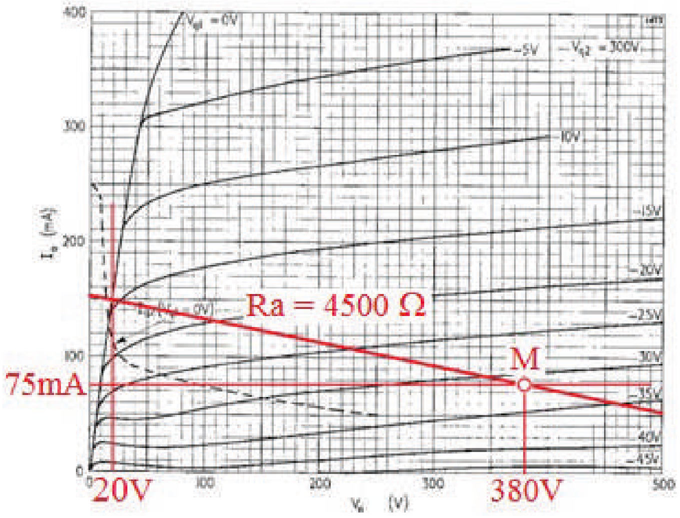




The characteristics

$I_a = f(U_a)$ of power beam tetrodes are more sharply curved (knee) than the characteristics of classical power pentodes. As a consequence of such characteristics shape, the harmonics distribution of the anode current is different from that of the pentodes. If the anode load line is placed to intersects the $U_{g1} = 0V$ characteristic at its knee, the third harmonic of the anode current is very small and the signal distortion is mainly caused by the second harmonic of the anode current.

Example:
Power beam tetrode KT88



Anode voltage (chosen): U_a *= 380 V*
Anode dissipation (chosen):
$P_d \le 0.7 \times P_{d\,max}$,
$P_d \le 0.7 \times 42\ W$, $P_d \le 29\ W$

Anode quiescent current:
$I_a = P_d / U_a = 29\ W / 380\ V = 76\ mA$
$I_a \approx 75\ mA$
Quiescent point M (380 V, 75 mA)
Using $I_a = f(U_a) \rightarrow U_{gk} = -32\ V$
Bias voltage: $U_{gk} = -32\ V$
Screen grid current: $I_{g2} \approx 3\ mA$

Calculation of anode load:

$$R_{Load} = 0.9 \times \frac{U_a}{I_a} = 0.9 \times \frac{380\ V}{0.075\ A}$$
$$R_{Load} = 4500\ \Omega$$

Draw Anode Load line R_{Load} = 4500 Ω through the quiescent point M (380 V, 75m A)

Calculation of R_k :

$R_k = U_{gk} / (I_a + I_{g2}) = 32V / (75\ mA + 3\ mA) = \mathbf{410\ \Omega}$

The anode circuit power supply voltage is:

$$U_{ab} = U_k + U_{ak} + U_{OT-DC}$$

$$U_{ab} = 32 + 380 + 8 = 420\,V$$

$^*U_{OT\text{-}DC}$ *is approximately 8V*

It is necessary to adjust the screen grid power supply voltage to a value of ***332 V****.*

$$R_{g2\ total} = R_{g2} = \frac{U_{ab} - 382}{I_{g2}} = \frac{420\ \mathrm{V} - 332\ \mathrm{V}}{3\ \mathrm{mA}} = \mathbf{29333\ \Omega}$$

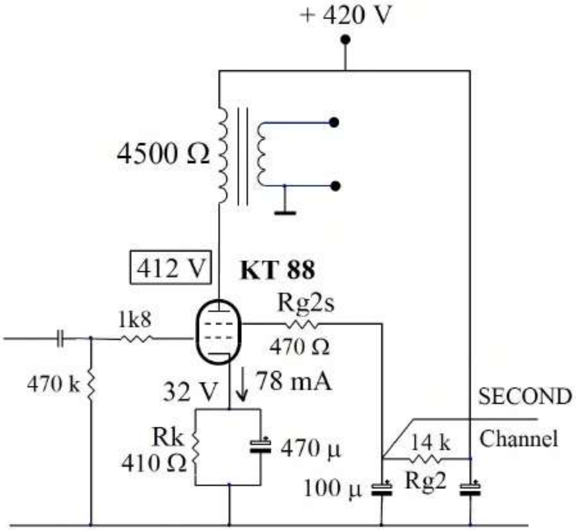
R_{g2} = 14 kΩ and two separate screen grid stoppers R_{g2S1} = R_{g2S2} = 470 Ω

**common power supply circuit is used for both channels of the amplifier*

Calculation of output power:

$$P_{out} = \frac{1}{2} \times (U_{ab} - U_r) \times I_a = \frac{1}{2} \times (380\ V - 20\ V) \times 0.075\ A = \mathbf{13.5\ W}$$

Or $P_{out} = \frac{1}{2} \times \frac{(U_a - U_r)^2}{R_{Load}} = \frac{1}{2} \times \frac{(380\ V - 20\ V)^2}{4500\ \Omega} = \mathbf{14.4\ W}$



4.11 PUSH-PULL Output stage configuration

*** Ideal Push-Pull output stage**

- The electrical characteristics of both tubes are equal
- Both tubes operate under the same operating conditions
- Output tubes are driven by signals of equal amplitudes but opposite phases
- Center-tap output transformer (i.e. half of the primary winding is an integral part of the circuit of each tube)

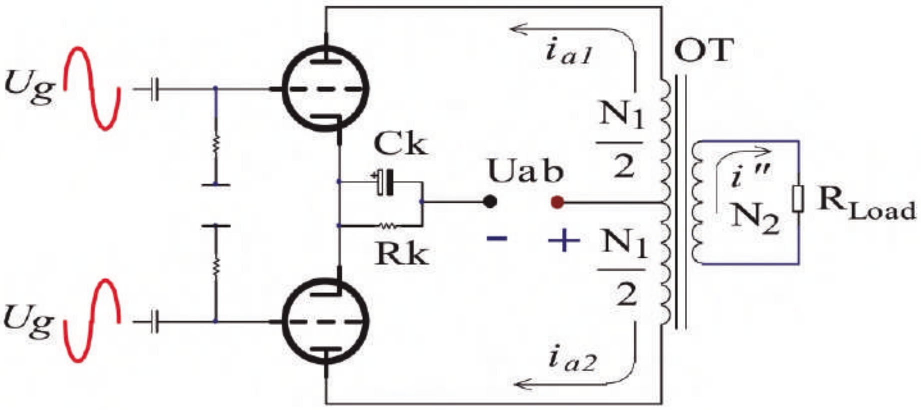


Fig. 4-54

Equivalent circuit:

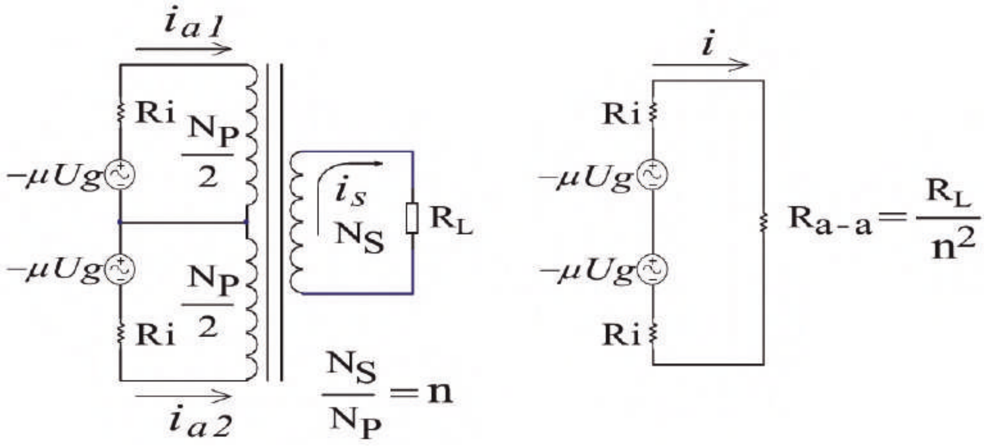


Fig. 4-55

The main characteristics of the Push-Pull output stage

- **Even harmonics** of the amplified signal are **canceled** (lower total harmonic distortion)

  For Fig. 4-55: $\frac{N_P}{2} \times i_{a1} - \frac{N_P}{2} \times i_{a2} = N_S \times i_S$

  Driving signal of the control grid of the first tube: $U_{g1} = U_{gm} \times \cos(\omega t)$

  Driving signal of the control grid of the second tube (signal of the opposite phase to the phase of the signal of the first tube):

  $U_{g2} = -U_{gm} \times \cos(\omega t) = U_{gm} \times \cos(\omega t + \pi)$

  Anode currents (with harmonics):

  $i_{a1} = I_a + I_{1m} \times \cos(\omega t) + I_{2m} \times \cos(2\omega t) + I_{3m} \times \cos(3\omega t) + \cdots$

  $i_{a2} = I_a + I_{1m} \times \cos(\omega t + \pi) + I_{2m} \times \cos 2(\omega t + \pi) + I_{3m} \times \cos 3(\omega t + \pi) + \cdots$

  Secondary winding current of the output transformer:

  $i_S = \frac{1}{2} \times \frac{N_P}{N_S} \times (i_{a1} - i_{a2}) = \frac{1}{2 \times n} \times (2I_{1m} \cos \omega t + 2I_{3m} \cos 3\omega t + \cdots)$ → The output signal contains only **odd harmonics: ωt, 3ωt, 5ωt, ...**

  The effect of canceling the even harmonics as well as the fact that the amplitudes of the harmonics decrease in order of appearance, is that the non-linear distortion of the push-pull stage is very low.

- **Non-pre-magnetization of the output transformer core caused by the opposite flow of DC current through each half of the primary winding.**

- **Canceling the unwanted effects of insufficient filtering of the DC power supply voltage as an effect of the symmetry of the Push-Pull circuit.**

Output power of the Push-Pull stage:

Amplitudes of anode voltage and current of one tube are:

$I_{1m} = \frac{N_S}{N_P} \times I''_m = n \times I''_m$, and $U_{1m} = \frac{1}{2} \times \frac{N_P}{N_S} \times U''_m = \frac{1}{2} \times \frac{1}{n} \times U''_m$

U_m'' and I_m'' - amplitudes of the fundamental voltage and current harmonic of the output transformer secondary.
The AC resistance R_a of the anode circuit of one tube (anode load) is:

$$R_a = \frac{U_{1m}}{I_{1m}} = \frac{\frac{1}{2}\times\frac{1}{n}\times U_m''}{n\times I_m''} = \frac{1}{2}\times\frac{1}{n^2}\times R_L$$

and the total AC resistance (load) between the anodes of the tubes (anode to anode load):

$$R_{aa} = \frac{1}{n^2}\times R_L$$

Using the equivalent circuit:

$$I_{1m} = \frac{\mu\times U_{gm}}{R_i+\frac{R_L}{2\times n^2}} \text{ and } U_{1m} = \frac{\mu\times U_{gm}}{R_i+\frac{R_L}{2\times n^2}}\times\frac{R_L}{2\times n^2},$$ (U_{gm} - amplitude of the driving voltage (signal))

Output power:

$$P_{out} = \frac{(\mu\times U_{gm})^2}{\left(R_i+\frac{R_L}{2\times n^2}\right)^2}\times\frac{R_L}{2\times n^2} \text{ or } P_{out} = \frac{1}{2}\times\frac{(\mu\times U_{gm})^2}{\left(\frac{R_i}{2}+\frac{R_{aa}}{4}\right)^2}\times\frac{R_{aa}}{4} \text{ or,}$$

$$P_{out} = 2\times\frac{(\mu\times U_{gm})^2}{(2\times R_i+R_{aa})^2}\times R_{aa}$$

P_{out} – output power
μ - tube amplification
R_i – internal resistance of the tube (plate (anode) resistance)
R_{aa} – anode (plate) to anode (plate) load
U_{gm} – maximum driving voltage (amplitude)

Output Transformer - The relation between the turns ratio and impedances of the transformer (transformer theory):

$$n = \frac{N_S}{N_P} = \sqrt{\frac{R_L}{R_{aa}}} \quad \text{or} \quad n^2 = \frac{R_L}{R_{aa}}$$

n – turns ratio of the push-pull output transformer
N_P – the total number of turns of the primary winding of the output transformer
N_S – the number of turns of the secondary winding of the output transformer
R_L – output transformer secondary load (AC resistance)
R_{aa} – anode to anode (plate to plate) load (AC resistance)

- **The optimal ratio R_i / R_{aa} is very important – it determines the maximum useful output power**

  The condition for optimal matching of a power generator with internal resistance R_i to the load R_L with which it is terminated (the generator delivers maximum power to the load) is: $R_i = R_L$
  If the previous applied to a push-pull output stage:

  R_{aa} / 4 = R_i / 2 (i.e. **R_i = R_a**).

A Push-pull output stage can be designed and built using triodes, tetrodes, or pentodes and tetrodes connected as triodes (triode mode of operation).

4.11.1 TRIODE CLASS A PUSH-PULL Output stage GRAPHICAL ANALYSIS

Note:
*It is not possible to analyze the push-pull amplifier configuration based on the characteristics of a single tube, because both tubes operate (conduct current) at the same time and influence each other. Therefore, it is necessary to design the **equivalent (composite) characteristics** of the push-pull stage configuration by defining the relations between anode voltage and anode current for both tubes at the same time.*

 # Valid for push-pull class A and class AB output stage configuration

Drawing the equivalent (composite) characteristics and the equivalent (composite) load line:

1. Set of the characteristics $I_a = f(U_a) | U_g = \text{const.}$
2. Operating (quiescent) point of tube M (350 V, 80 mA) $| U_g = -70$ V
3. Another set of the characteristics $I_a = f(U_a) | U_g = \text{const.}$ rotate by 180^0 and position them so that the U_a axis touches the U_a axis of the non-rotated set of characteristics and adjust their position so that the operating voltages of the M points align vertically.
4. **Equivalent (composite) characteristics**
   The consequence of the fact that the tubes of the push-pull configuration are driven by signals of equal voltage amplitudes but opposite in phase is that the increase of the anode voltage of one tube is equal to the decrease of the anode voltage of the other tube at any moment.
   The current of the secondary of the output transformer is proportional to the difference of the currents flowing through the halves of the primary winding at any moment.
   Equivalent characteristics represent the difference between the anode currents of the first and second tubes as a function of the anode voltage for a different voltage of the control grid:

   $$(i_{a1} - i_{a2}) = f(u_a) \mid u_g = const.$$

   Draw a vertical line $U_a = 350$ V. It intersects the upper and lower characteristic $U_g = -70$ V at equal $I_a = 80$ mA. Draw the first point of the composite characteristic at $U_a = 350$ V and $I_a = 0$ ma (80 mA – 80 mA = 0 mA). Draw the second vertical line $U_a = 370$ V. It intersects the upper characteristic $U_g = -70$ V at $I_a = 125$ mA and the lower characteristic $U_g = -70$ V at $I_a = 45$ mA. Draw the second point of the composite characteristic at $U_a = 370$ V and $I_a = (125 - 45)$ mA = 80 mA. Repeat the above process for a few more points and draw a line through them. It is the first composite characteristic. This line also defines the internal resistance R_{ieq} of the composite tube. Process the second pair of characteristic: upper $U_g = -80$ V and lower $U_g = -60$ V, in the same way as above (drawing vertical lines for some values of U_a, reading the value of Ia at the intersection of the vertical lines of the upper and lower characteristic, calculate the equivalent value) and draw the second composite characteristic.

5. Do the same for other pairs of the characteristics: (- 90 V, - 50 V), (- 100 V, - 40 V), (- 110 V, - 30V), (- 120 V, - 20V), (- 130 V, - 10 V), (- 140 V, 0V).

6. Draw the load line of the upper and lower tube as it is drawn in the case of single ended amplifier.

7. **Equivalent (or Composite) Load Line ($R_{a\ eq}$)**
   Draw a straight line through the points A and A located at the intersections of $I_a = I_{a\ min\ and}$ $I_a = I_{amax}$ and the new composite characteristics $I_a = f(U_a) | U_g = 0$ V and $U_g = -140$ V, respectively. This is so-called **Equivalent (or Composite) Load Line**.

8. The slope of the Equivalent (or Composite) Load Line is the impedance seen by one tube, and it is equal to 1 / 4 of the actual anode-to-anode load impedance.

   $$\boldsymbol{R_{a-a} = 4 \times R_{a\,eq}}$$

9. The above analysis is shown in Fig. 4-56 (2 x 300B triodes push-pull output stage).

Calculation of R_{aeq}, R_{a-a}, P_{out}

- R_{aeq} is the slope of the composite load line, i.e. $U_a / \Delta I_a = (700 - 0) / (0.256 + 0.256) = 1367\ \Omega$.

- $R_{aeq} = R_{a-a} / 4 \rightarrow R_{a-a} = 4 \times 1367 = 5468\ \Omega$

- $\boldsymbol{P_{out}} = \frac{(U_{a\,max} - U_{a\,min}) \times (I_{a\,max} - I_{a\,min})}{8} = \frac{(568.75 - 133.75) \times (0.158 + 0.158)}{8} = \mathbf{17.18\ W}$

- Or: $\boldsymbol{P_{out}} = \frac{U^2}{R} = \frac{\left(\frac{U_a - U_{a\,min}}{\sqrt{2}}\right)^2}{R_{a\,eq}} = \frac{\left(\frac{350 - 133.75}{\sqrt{2}}\right)^2}{1367} = \mathbf{17.2\ W}$

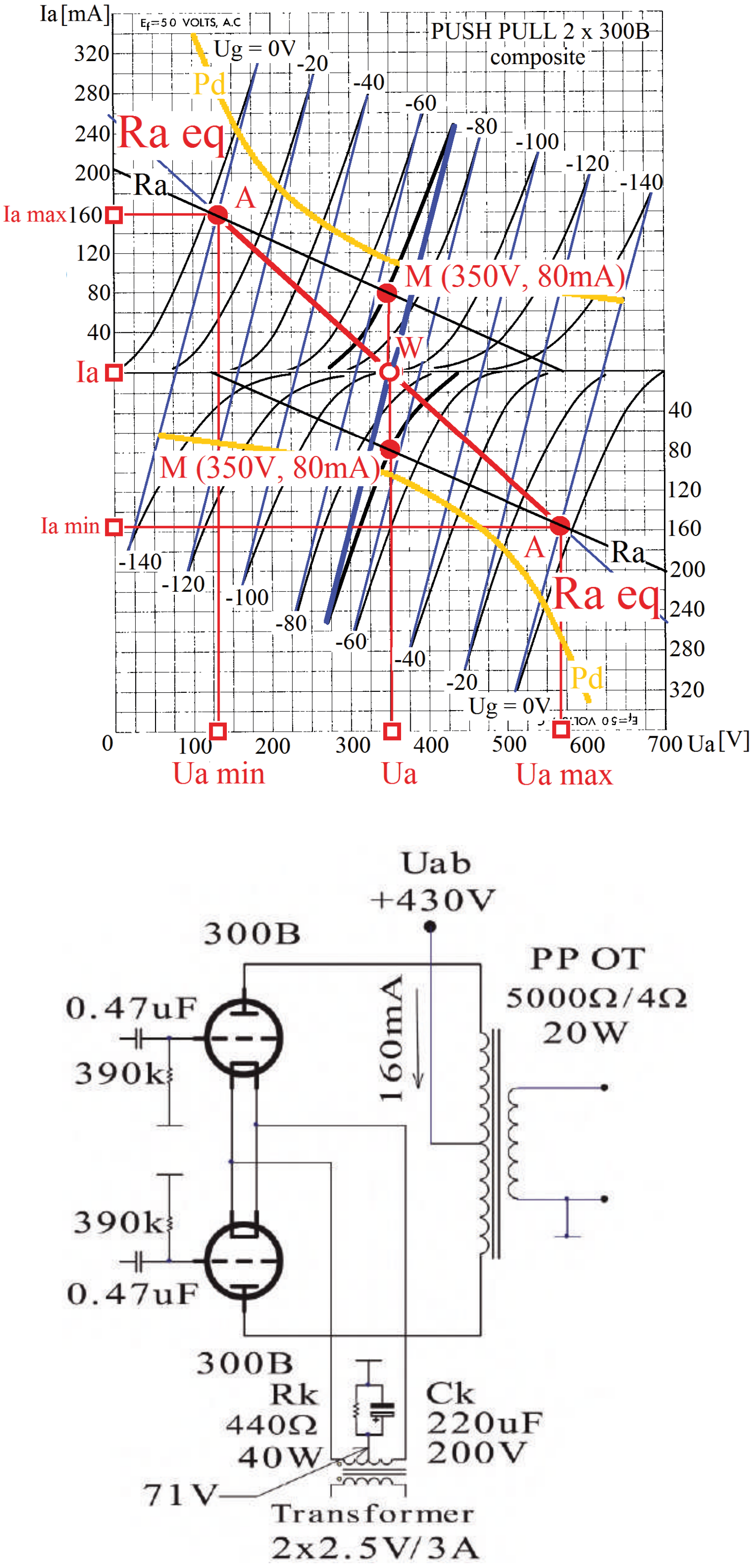
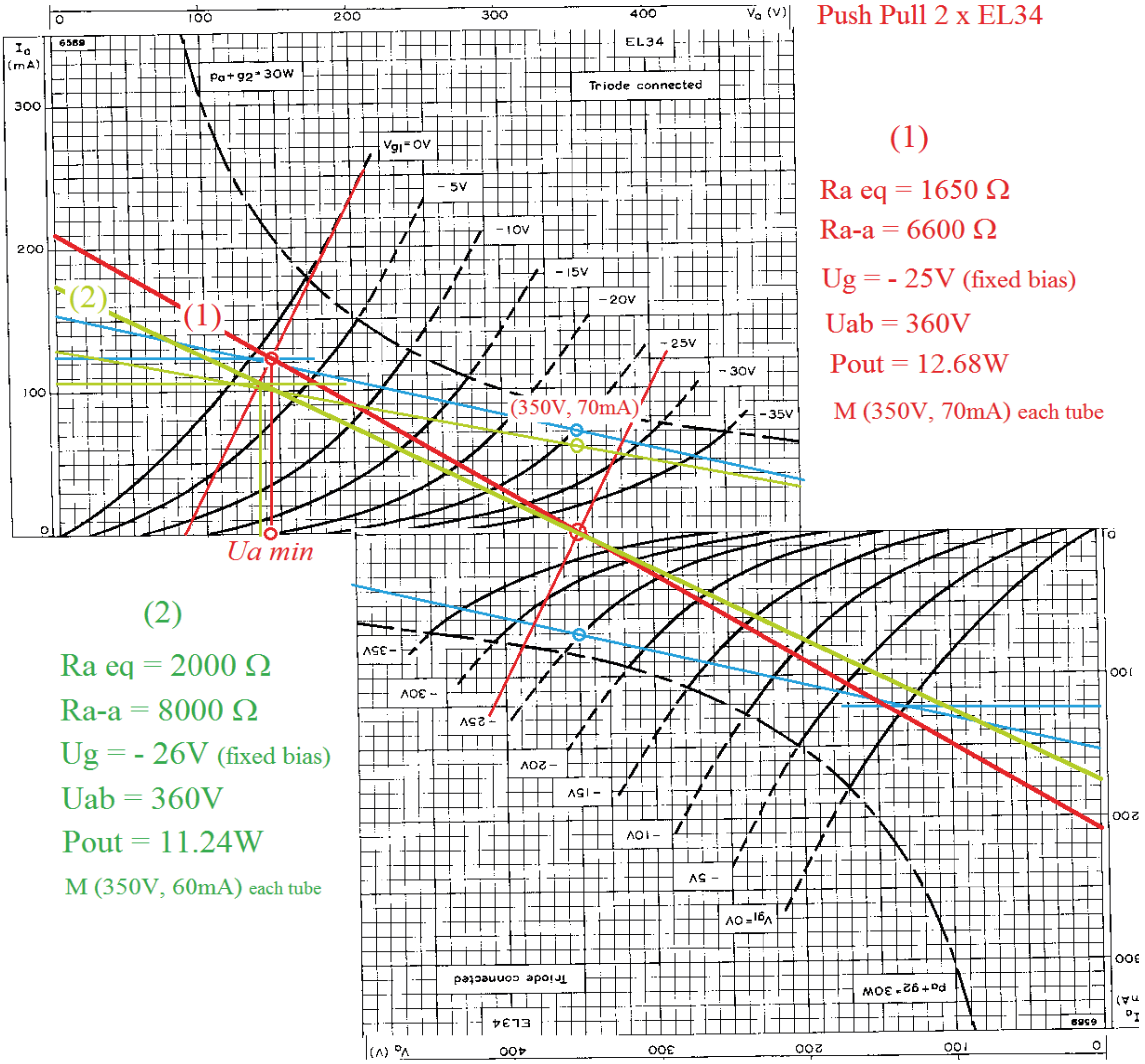

Ia [mA]
Ef=5.0 VOLTS, A.C
Ug = 0V
-20
-40
-60
-80
-100
-120
-140
PUSH PULL 2 x 300B
composite
Pd
Ra eq
Ra
A
Ia max
M (350V, 80mA)
W
Ia
Ia min
320
280
240
200
160
120
80
40
0
100
200
300
400
500
600
700 Ua [V]
Ua min
Ua
Ua max
Uab
+430V
300B
0.47uF
390k
160mA
PP OT
5000Ω/4Ω
20W
Rk
440Ω
40W
Ck
220uF
200V
71V
Transformer
2x2.5V/3A

Fig. 4-56

Example: Push Pull 2 x EL 34 (triode mode - connected as a triode)



Push Pull 2 x EL34

(1)

Ra eq = 1650 Ω

Ra-a = 6600 Ω

Ug = - 25V (fixed bias)

Uab = 360V

Pout = 12.68W

M (350V, 70mA) each tube

(2)

Ra eq = 2000 Ω

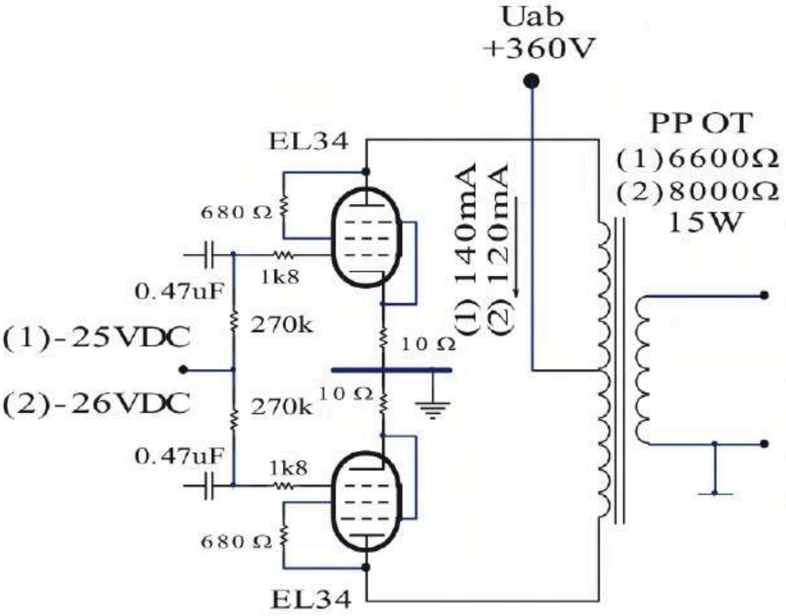
Ra-a = 8000 Ω

Ug = - 26V (fixed bias)

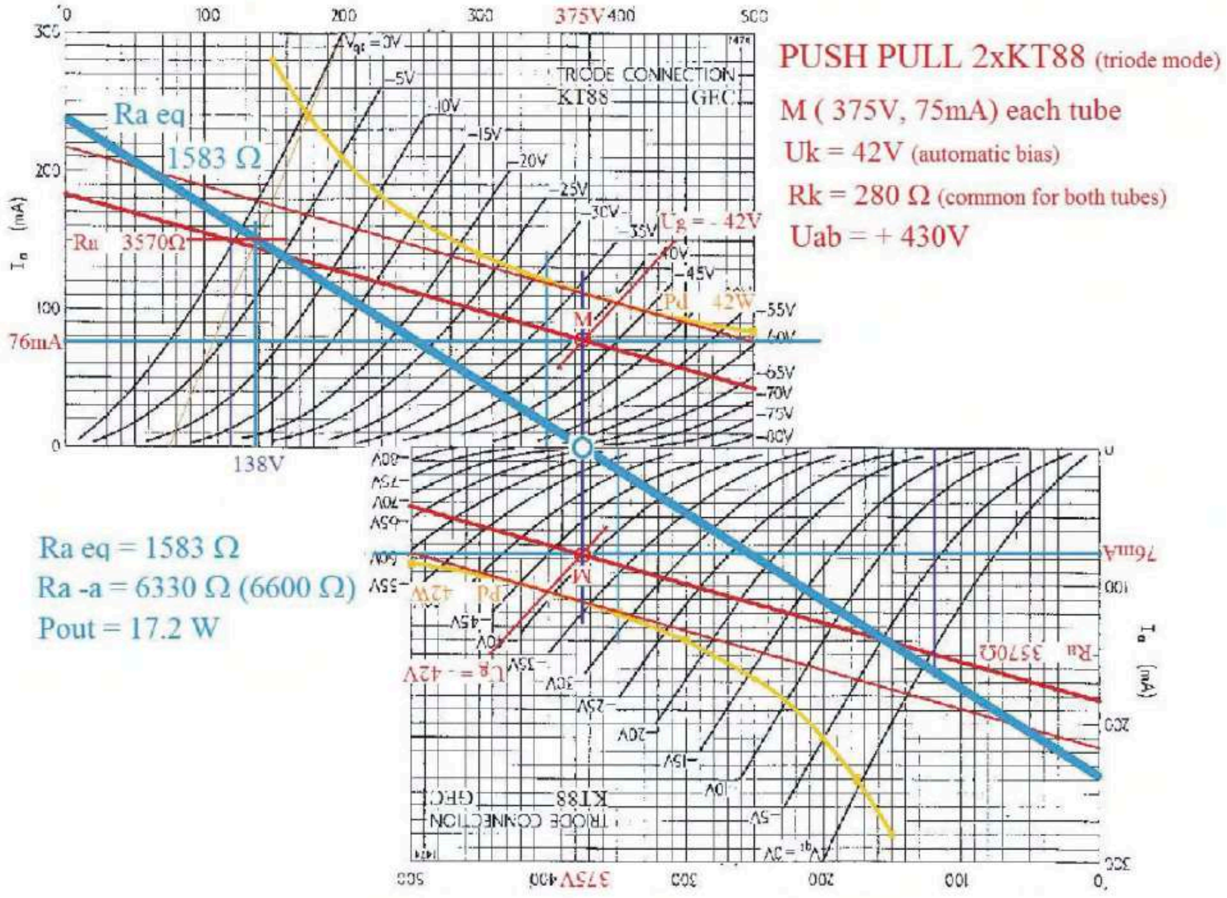
Uab = 360V

Pout = 11.24W

M (350V, 60mA) each tube



Example: Push Pull 2 x KT 88 (triode mode - connected as a triode)



4.11.2 PENTODE and BEAM TETRODE PUSH-PULL Output stage

The pentode and beam tetrode push-pull output stage can be designed and analyzed based on the basic principles applied in design and analysis of the triode PP output stage.

 # Pentodes and beam tetrodes have a screen grids. So, it is necessary to take care of the power supply of the screen grid. It can be built using the separate power supply unit or using the power supply of the anode circuit (resistor voltage divider).

The methodology of graphical analysis of triode PP circuit can be applied to the graphical analysis of pentode and beam tetrode PP circuit.

Drawing the equivalent (composite) load line:

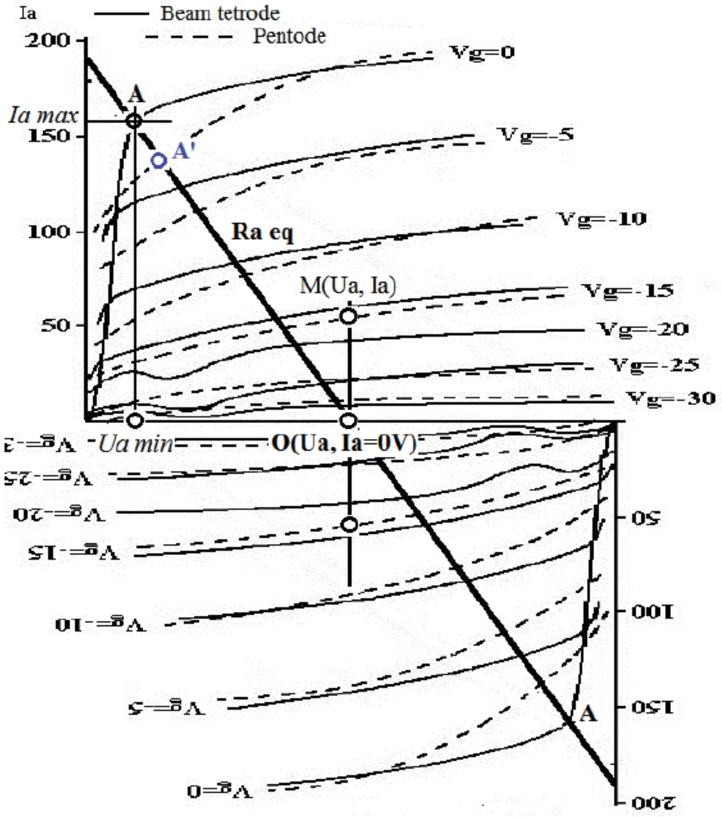
1. Set of the characteristics $I_a = f(U_a) | U_g = \text{const.}$
2. Operating (quiescent) point of tube M (U_a, I_a) $| U_g = \text{const}$
3. Another set of the characteristics $I_a = f(U_a) | U_g = \text{const.}$ rotate by 180^0 and position them so that the U_a axis touches the U_a axis of the non-rotated set of the characteristics and adjust their position so that the operating voltages of the M points align vertically.
4. Equivalent (or Composite) Load Line ($R_{a\ eq}$)

   Draw a straight line through the point 0 (Ua, Ia = 0) and the knees (curvature of the characteristic) of the characteristic U_g = 0V of both tubes.

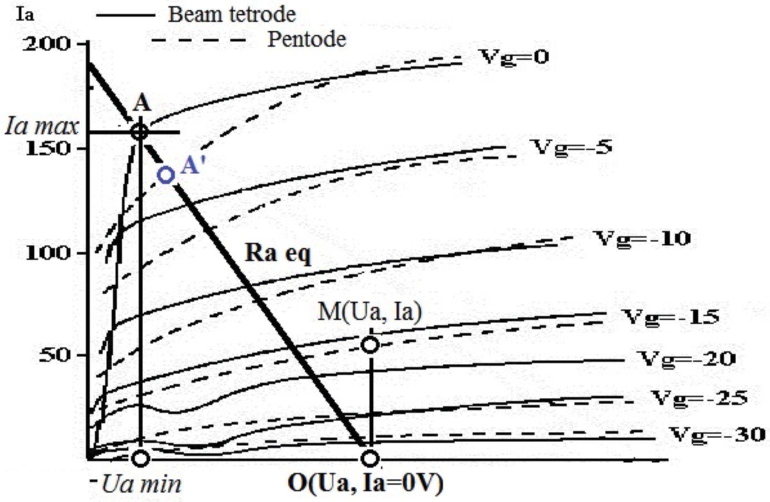
   This is the so-called **Equivalent (or Composite) Load Line**.

5. The slope of the Equivalent (or Composite) Load Line is the impedance seen by one tube ($R_{a\,eq} = \frac{U_a - U_{a\,min}}{I_{a\,max}}$), and it is equal to 1 / 4 of the actual anode-to-anode load impedance.

$$R_{a-a} = 4 \times R_{a\,eq}$$



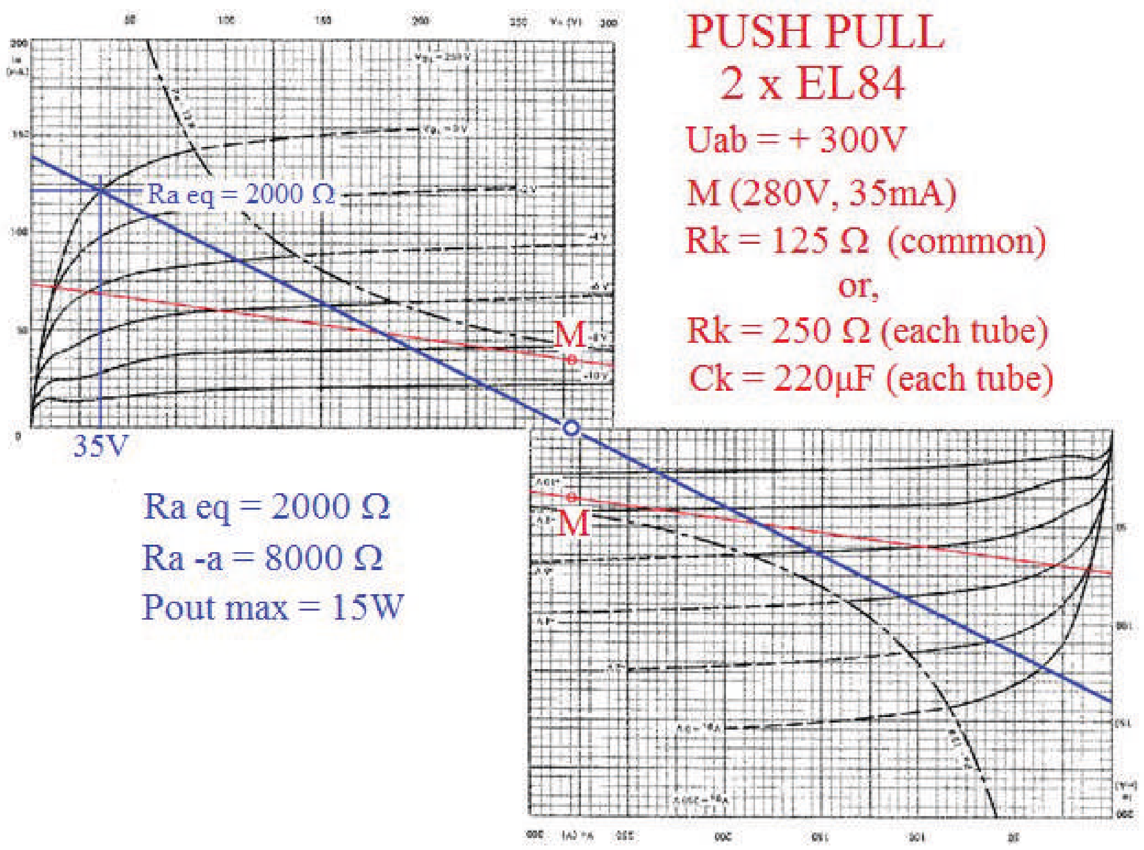
By analyzing the above graphic, it can be concluded that it is not necessary to draw both characteristics to define the equivalent load line. It is enough to use one characteristic and draw a straight line through the points O (Ua, Ia = 0) and the knee of the characteristic Ug1 = 0V and calculate the slope of the line (equivalent load line).

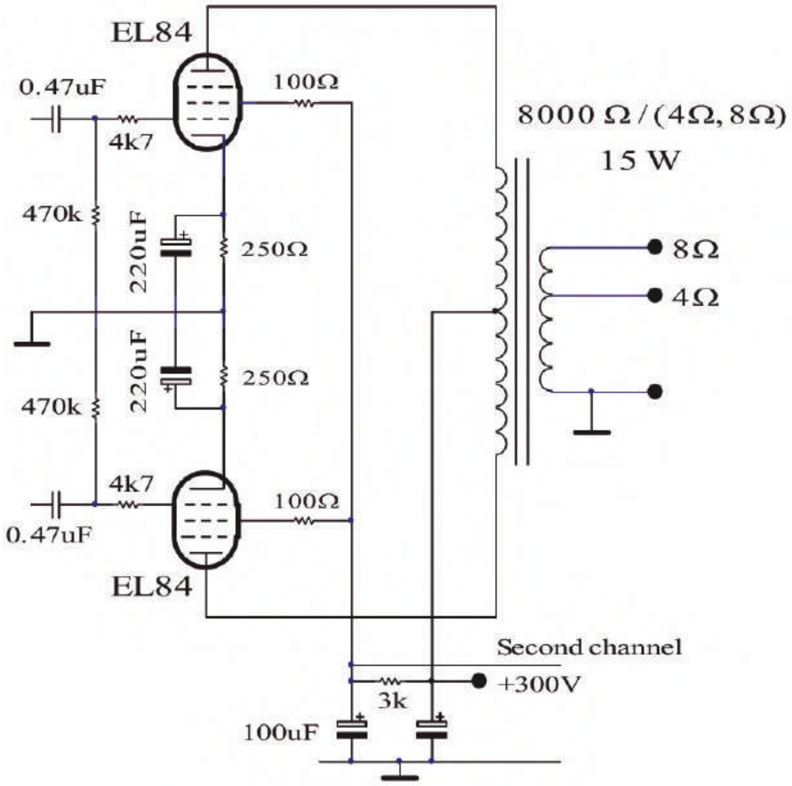


Output power:

$$P_{out} = \frac{1}{2} \times I_{a\,max} \times (U_a - U_{a\,min})$$

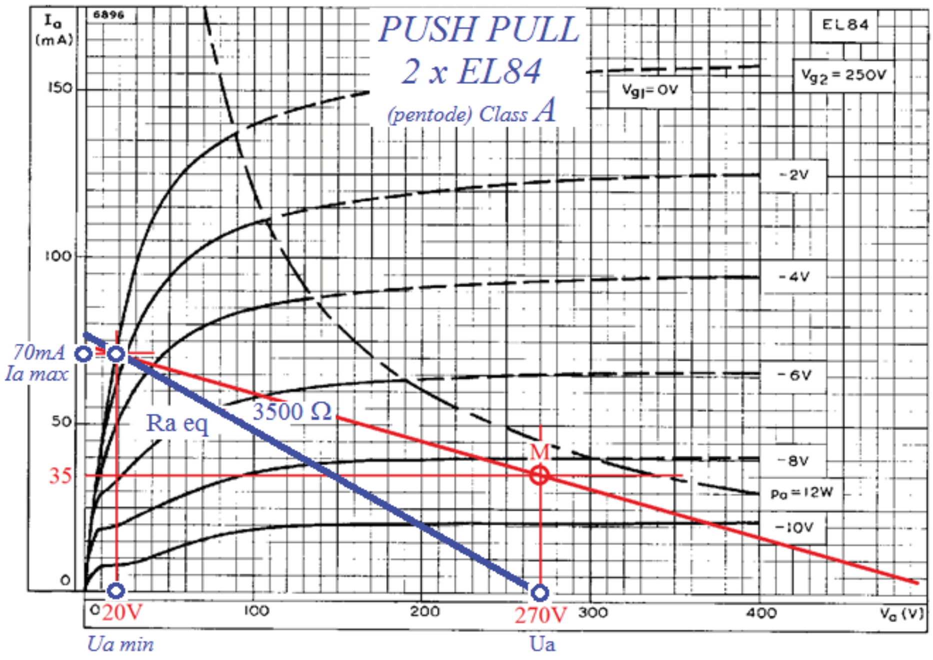
Example:
Push Pull 2 x EL84
M (280 V, 35 mA)





Push-Pull 2 x EL84 Class A (pentode):

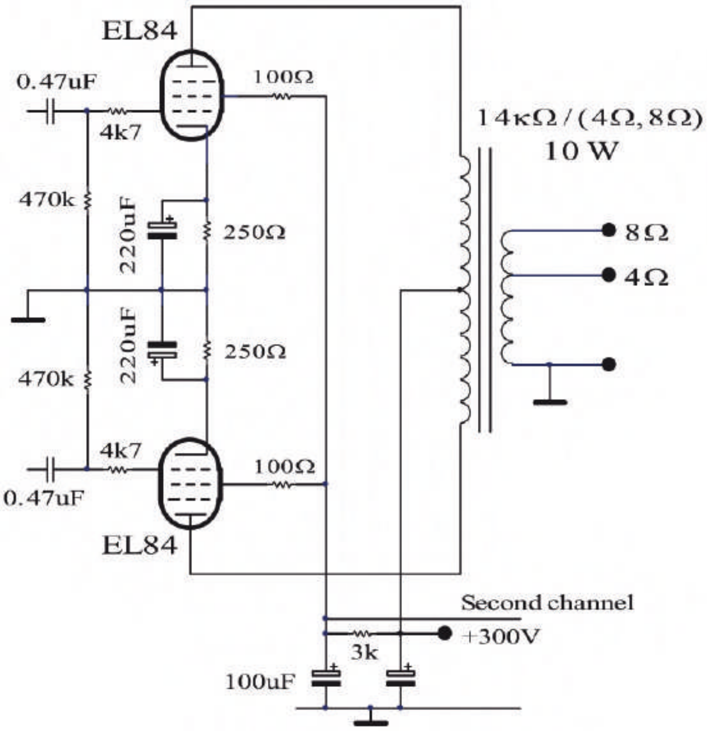
Graphical analysis
Draw the SE class A EL 84 load line on the characteristic diagram.
Draw a straight line through the point (Ua, Ia = 0) and the intersection point of the characteristic $U_{g1} = 0V$ *and the SE load line.*
This line is the equivalent push-pull load line: $R_{a\ eq} = R_{a\ a\ (push\text{-}pull)} / 4$



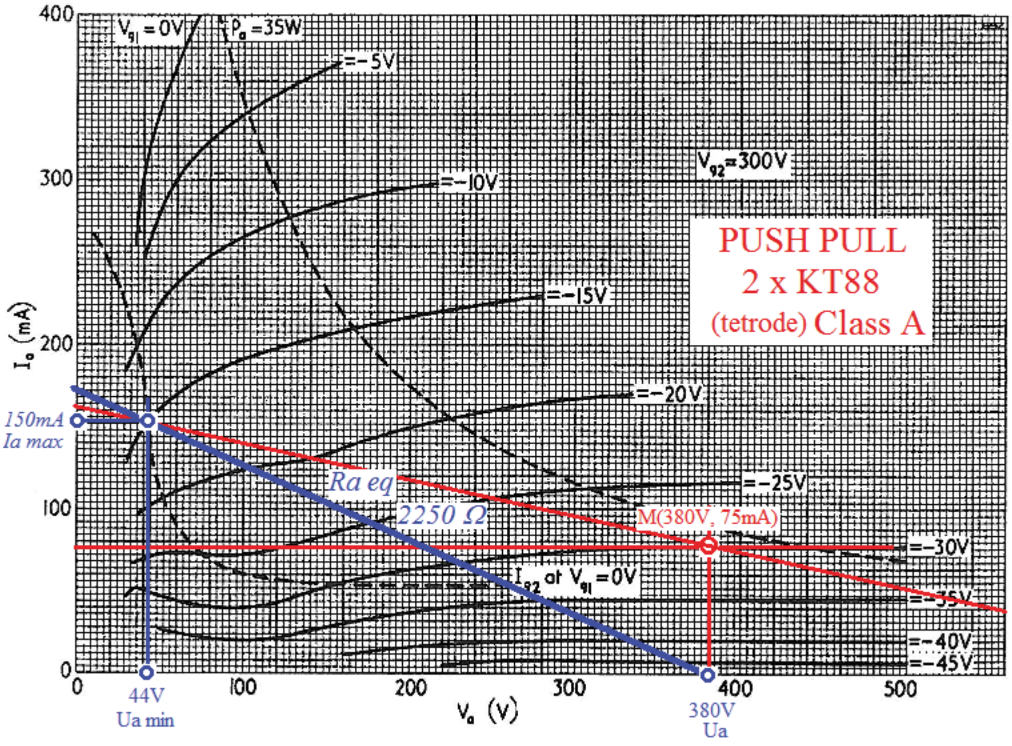
Using the diagram:

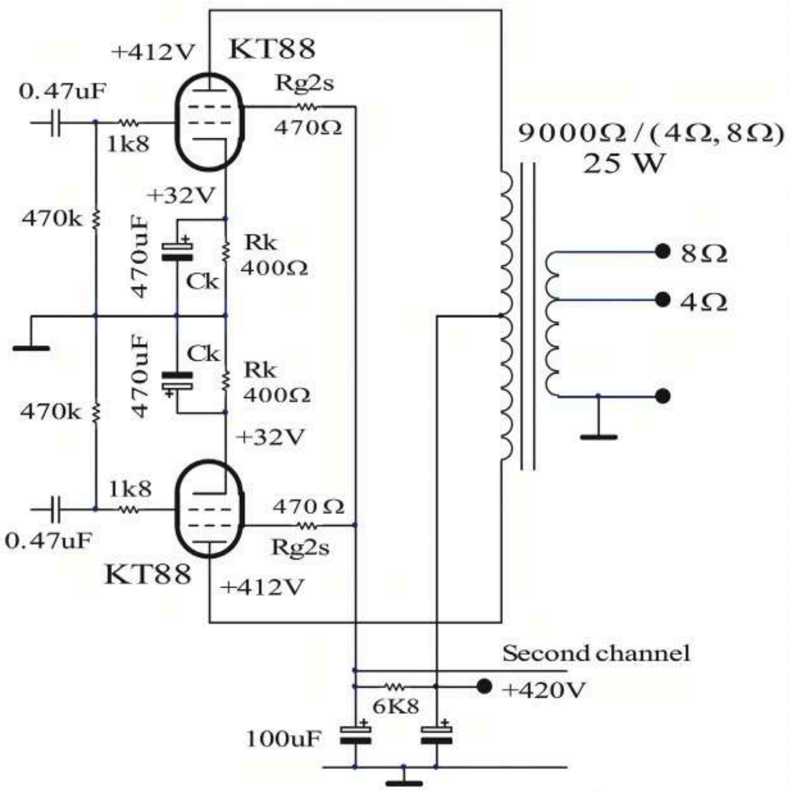
$$R_{a\,a} = 4 \times R_{a\ eq} = 4 \times 3500\ \Omega = \mathbf{14\ k\Omega}$$

$$P_{out} = 2 \times \frac{(U_a - U_{a\,min})^2}{R_{a\,a}} = 2 \times \frac{(270-20)^2}{14000} = 8.9\,W$$



Example:
Push-Pull 2 x KT88
M (380 V, 75 mA)



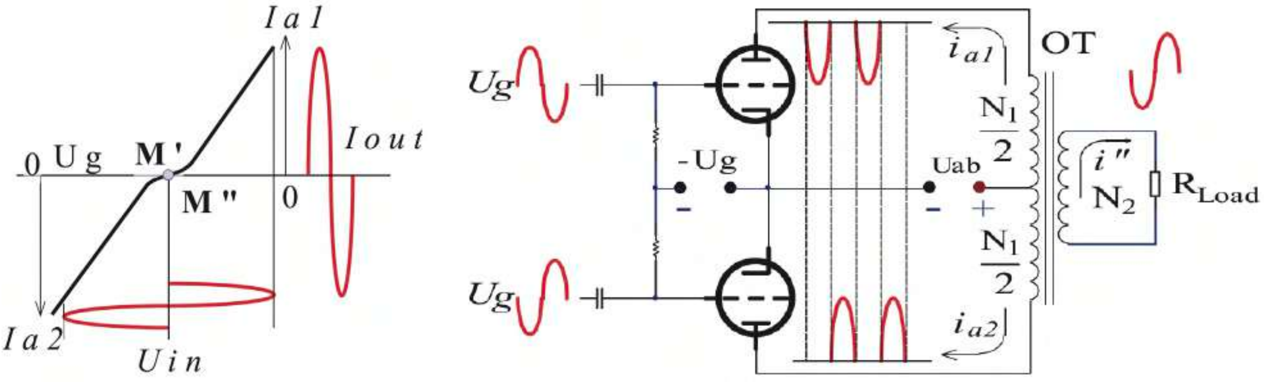


4.11.3 CLASS-AB PUSH-PULL Output stage

Class-B push-pull amplifier

Quiescent point is located on the lower knee of the characteristic $I_a = f(U_{g-k})$ *i.e. the control grid bias voltage corresponds to the anode current cut-off (*$I_a = 0$*). In the absence of a driving signal on the control grid, the anode current doesn't flow. If a sine wave signal is applied to the control grid, the anode current flows only during the positive half cycle of the input signal. During the negative half cycle of the input signal, the tube doesn't conduct current. The total harmonic distortion of class B amplifier is high (high odd harmonic distortion and distortion generated in the crossover region – at low output power THD is maximum), so class B is inapplicable in high end amplifier designs.*

Class B push-pull amplifier has maximum efficiency ($\eta = 78.5\%$ *for sine wave signal) [*in the audio amplifiers class].*
Only a fixed bias of the control grid must be applied.
Without going into details, some useful equations and conclusions that can be used in the design process of class B push-pull amplifiers:



- The slope of the composite characteristic is half the slope of the composite characteristic of a class A amplifier.

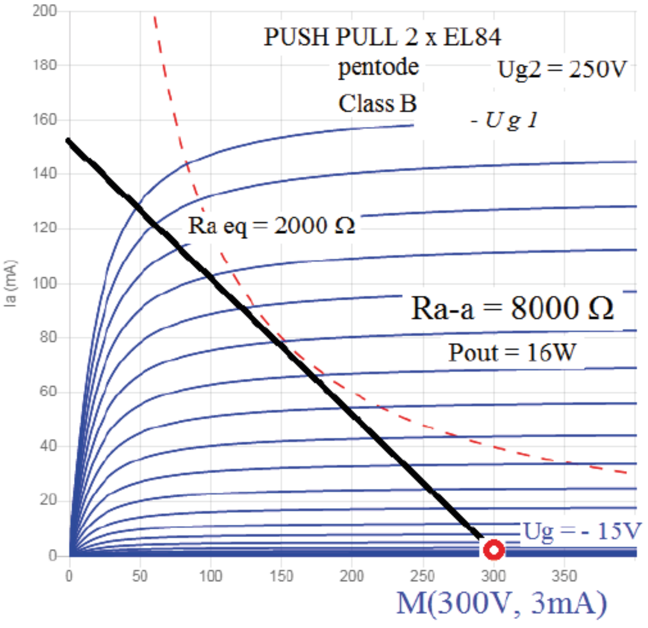
- The slope of the composite anode resistance is twice that of the composite anode resistance of class A amplifier.

Output power:

$$P_{out} = \frac{1}{2} \times I_{a\,max} \times (U_a \times U_{a\,min})$$

$$P_{out} = \frac{1}{8} \times I_{a\,max}^2 \times R_{a-a}$$

$$P_{out} = 2 \times \frac{(U_a - U_{a\,min})^2}{R_{a-a}}$$



Maximum power condition:

$$R_{a-a} = 4 \times R_i$$

Load resistance (anode to anode) $R_{a\text{-}a}$:

$$R_{a-a} = 4 \times \frac{U_a - U_{amin}}{I_{a\,max}}$$

Average anode current (each tube):

$$I_a = \frac{I_{amax}}{\pi} \approx 0.318 \times I_{amax}$$

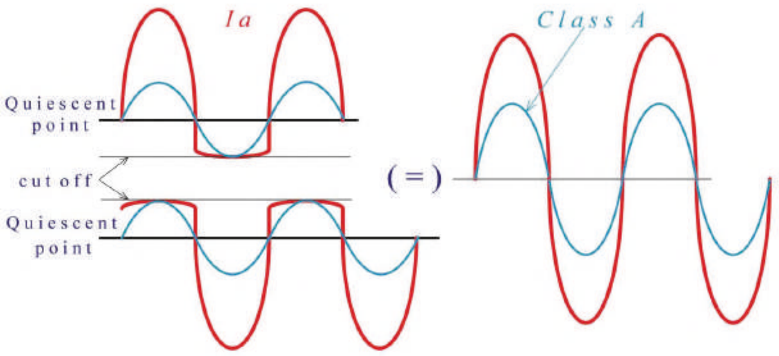
Anode circuit efficiency:

$$\eta = 0.785 \times \left(1 - \frac{U_{amin}}{U_a}\right)$$

Anode power dissipation:

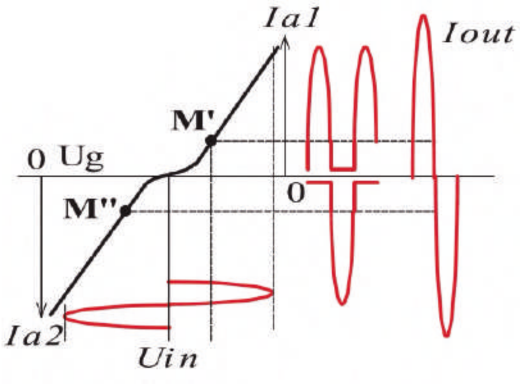
$$P_d = I_{a\,max} \times (0.137 \times U_a + 0.5 \times U_{a\,min})$$

CLASS-AB PUSH-PULL AMPLIFIER



The main disadvantage of class-B amplifiers: high THD, especially at low signal levels, can be partially eliminated by operation of the amplifier in class AB. The bias voltage of a tube operating in class AB is adjusted in such a way that for low drive signal the anode current flows for a full cycle of the drive signal. Under condition of higher drive voltage, anode current flows in each tube for appreciably more than one half of a cycle, but not for a complete cycle of the AC drive signal.

The tube bias can be designed as either fixed or automatic bias, but the amplifier design procedure is not the same for fixed and automatic bias since the DC anode currents of the tubes depend on and change with the driving AC signals.



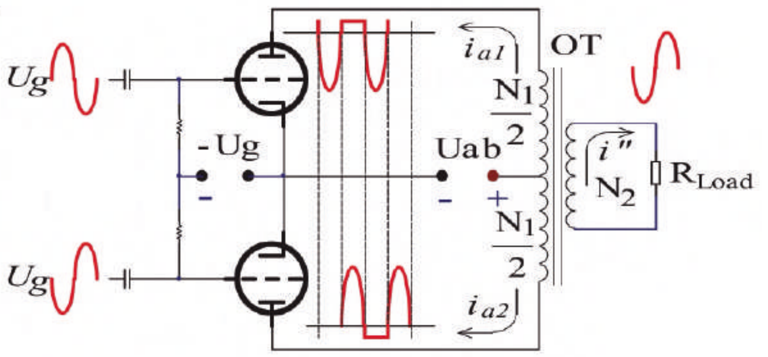
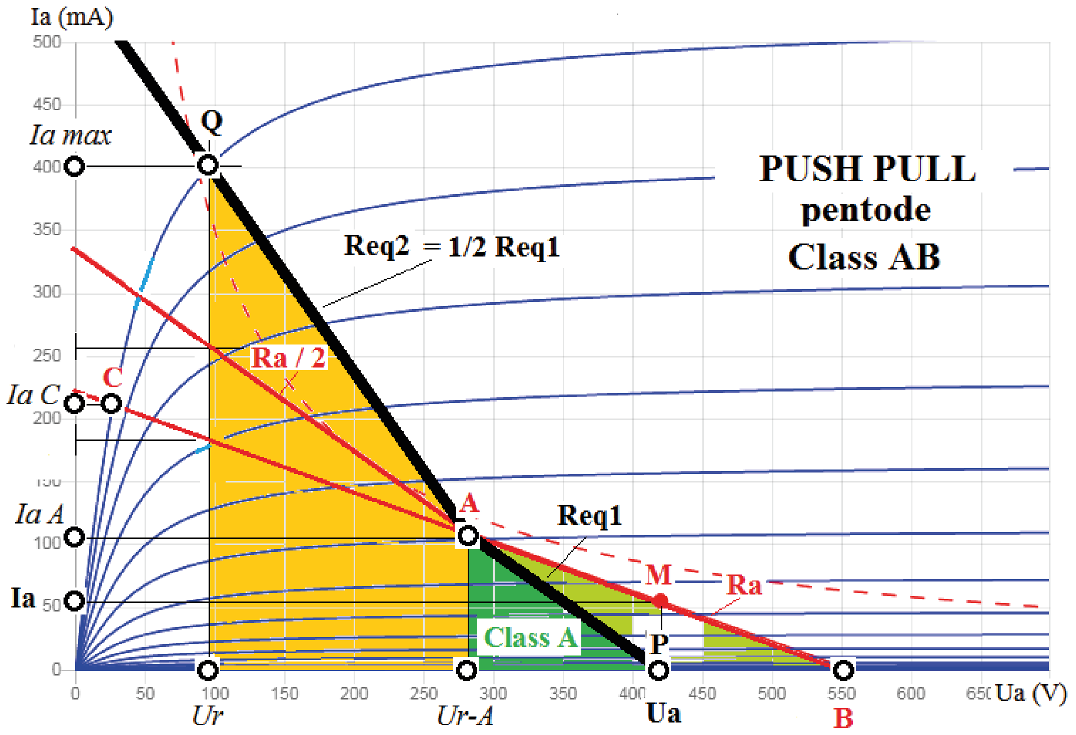


Illustration of a graphical analysis of a fixed-bias class AB pentode push-pull amplifier:



Analyzing the diagram these observations can be made:

For low drive signals (opposite in phase), the operating point M of one tube moves along the load line M-A, and at the same time, the operating point of the other tube moves along the load line M-B. At these values of the AC drive signal, both tubes conduct current and the operating conditions are adequate to the operating conditions of a class A amplifier.

Consequently, the anode load of each tube is: $R_a = \frac{R_i}{2 \times n^2}$

For a higher driving voltage than defined above, one of the tubes stops conducting current (at cut off point B) and current flows only through half of the output transformer primary. Consequently, the tube anode load changes to the value (to the left of point A): $\frac{R_i}{4 \times n^2} = \frac{R_a}{2}$ (two times lower than the load in the A-B region).

The graphical analysis is similar to the graphical analysis of the classical push-pull amplifier (draw the first equivalent load line R_{eq1} through the points P (U_a , $I_a = 0$) and A (intersection point with the single tube load line R_a). Draw the second equivalent load line $R_{eq2} = R_{eq1} / 2$ through point A to the intersection with the characteristic

$I_a = f(U_a) \mid U_{g1} = 0$ V)).

The total output power can be calculated by calculating the area of the geometric figure determined by the points: U_r – U_a (P) – A – Q –U_r:

$$P_{out}=\frac{\left[\left(U_a-U_{r-A}\right)\times I_{aA}\right]}{2}+\left(U_{r-A}-U_r\right)\times I_{aA}+\frac{\left[\left(U_{r-A}-U_r\right)\times\left(I_{a\,max}-I_{aA}\right)\right]}{2}$$

A class-AB push-pull amplifier operates in class A up to the output power:

$$P_{out\ class\ A}=\frac{\left[\left(U_a-U_{r-A}\right)\times I_{aA}\right]}{2}$$

The total output power can also be calculated using the equation:

$$P_{out}=\left(I_{a(C)}-I_a\right)^2\times\frac{R_{La-a}}{2}$$

** Above numerical analysis is based on the manufacturer's recommended application note for class AB push-pull 2 x KT88 tubes:*

U_a = + 420 V, I_a = 55 mA, U_{gk} = - 32.7 V, U_{g2} = +300 V, $R_{L\ a\text{-}a}$ = 5000 Ω

Using the graphical analysis diagram above:

U_{r-A} = 282.5 V; U_r = 99 V

I_{aA} = 110 mA; $I_{a(C)}$ = 205 mA; $I_{a\ max}$ = 404 mA

$R_{a\ eq1}$ = ($R_{L\ a\text{-}a}$ / 4) =5000 / 4 =1250 Ω; $R_{a\ eq2}$ = ($R_{a\ eq1}$ / 2) = 625 Ω

$$P_{out}=\frac{[(U_a-U_{r-A})\times I_{aA}]}{2}+(U_{r-A}-U_r)\times I_{aA}+\frac{[(U_{r-A}-U_r)\times(I_{a\,max}-I_{aA})]}{2}$$

$$=\frac{[(420-282.5)\times 0.11]}{2}+(282.5-99)\times 0.11+\frac{(282.5-99)\times(0.404-0.11)}{2}=54.7\,W$$

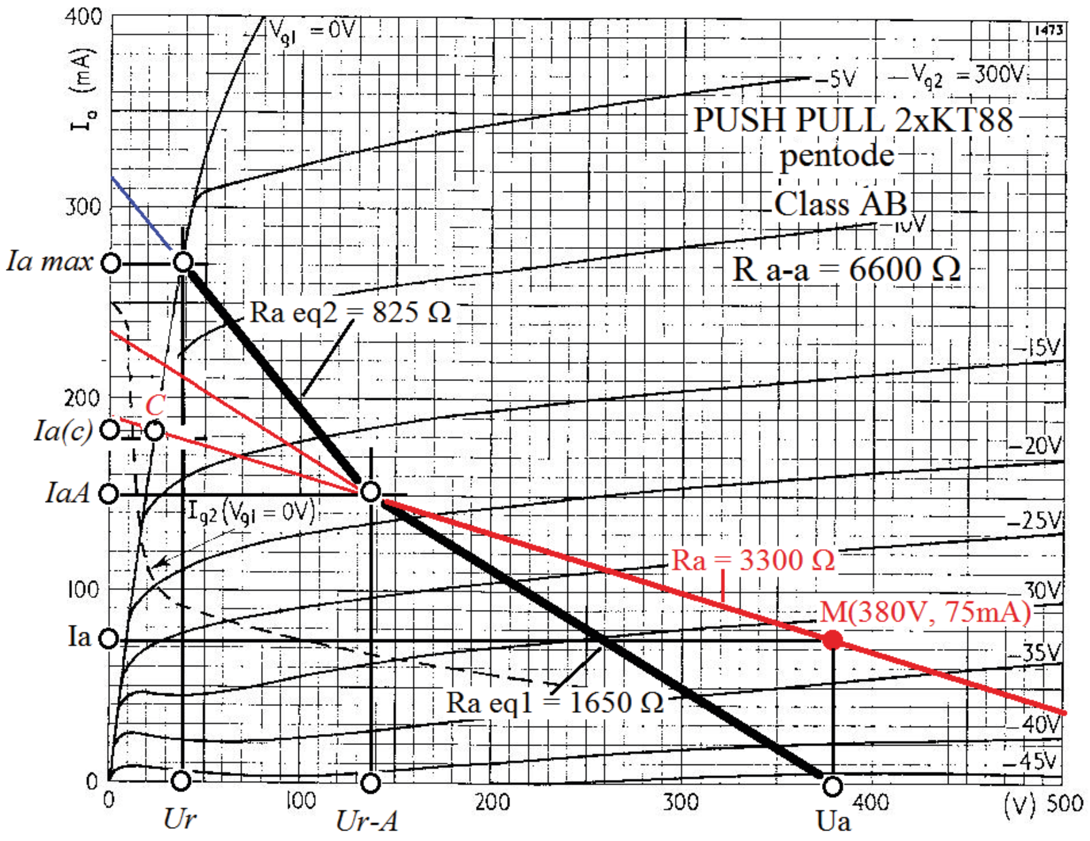
$$\boldsymbol{P_{out}}=\left(I_{a(C)}-I_a\right)^2\times\frac{R_{L\,a-a}}{2}=(0.205-0.055)^2\times\frac{5000}{2}=\boldsymbol{56.25\,W}$$

$$P_{out\ class\ A}=\frac{[(U_a-U_{r-A})\times I_{aA}]}{2}=\frac{[(420-282.5)\times 0.11]}{2}=7.56\,W$$

By moving the tube bias (quiescent point) closer to class A, the output power decreases, but the total harmonic distortion also decreases.

The price of lower total harmonic distortion is lower output power.

Example:
Class AB Push pull amplifier - 2 x KT88 pentodes



Using the diagram:

U_a = 380 V; U_{r-A} = 137 V; U_r = 39 V; I_a = 75 mA; $I_{a\,A}$ = 150 mA; $I_{a\,(C)}$ = 183 mA; $I_{a\,max}$ = 270 mA;
$R_{a\,eq1}$ = 1 / 4 $R_{a\text{-}a}$ = 1650 Ω; $R_{a\,eq2}$ = 1 / 2 $R_{a\,eq1}$ = 825 Ω; $R_{a\,-a}$ = 6600 Ω;

$$\boldsymbol{P_{out}} = \frac{[(U_a - U_{r-A}) \times I_{aA}]}{2} + (U_{r-A} - U_r) \times I_{aA} + \frac{[(U_{r-A} - U_r) \times (I_{a\,max} - I_{aA})]}{2}$$

$$= \frac{[(380 - 137) \times 0.15]}{2} + (137 - 39) \times 0.15 + \frac{(137 - 39) \times (0.27 - 0.15)}{2} = \mathbf{38.745\,W}$$

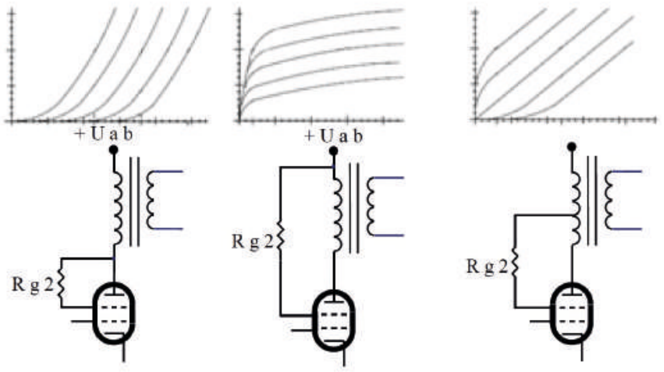
$$\boldsymbol{P_{out}} = \left(I_{a(C)} - I_a\right)^2 \times \frac{R_{L\,a-a}}{2} = (0.183 - 0.075)^2 \times \frac{6600}{2} = \mathbf{38.49\,W}$$

$$\boldsymbol{P_{out\,class\,A}} = \frac{[(U_a - U_{r-A}) \times I_{aA}]}{2} = \frac{[(380 - 137) \times 0.15]}{2} = \mathbf{18.225\,W}$$

The amplifier operates in class A up to 18.2W output power (almost 50% of nominal power (47%)), but the total distortion is very low (THD ≤ 0.2 %).

4.11.4 ULTRA LINEAR MODE PUSH-PULL Output Stage

In order to partially eliminate the disadvantages of the previously known output stage configurations (high THD of the pentode output stage and low output power of the triode output stage), Hafler and Keroes launched a specific tetrode (pentode) push-pull output stage configuration known as ultra – linear mode.

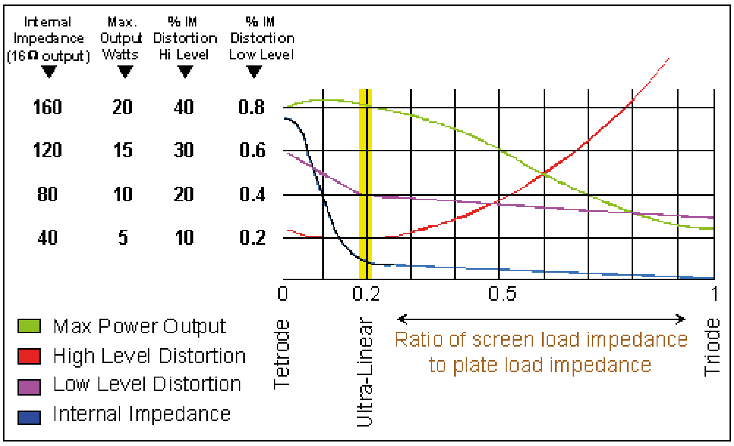


Tetrode (pentode) connected as a triode: the screen grid is connected to the anode of the tube.

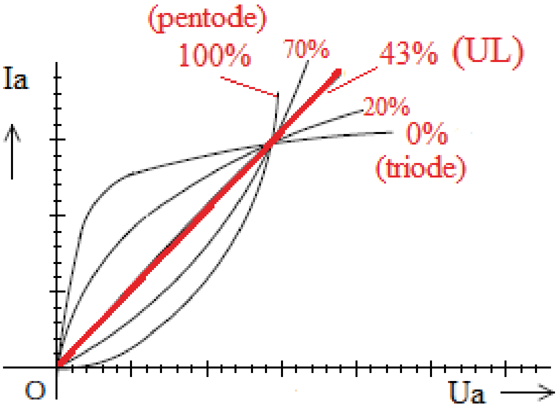
Base connection of tetrode (pentode): the screen grid is connected to the power supply.

Ultra linear mode: the screen grid (g_2) of tetrode (pentode) is connected to the tap (UL tap) of the primary winding of the output transformer located somewhere between the plate (anode) tap and the center tap of the primary winding. So, a portion of the plate (anode) signal is fed – back to the screen grid. By varying the magnitude of the fed – back signal to the screen grid (i.e. by varying the location of the UL tap), the operating mode of the tube changes from more-triode to more-tetrode (pentode). Moving the UL tap closer to the plate (anode) terminal of the primary of the output transformer makes the tube operate more like a triode, and moving the UL tap closer to the central tap (CT) of the primary windings makes the tube operate more like a tetrode (pentode).

The value of the optimal screen grid impedance is within a narrow impedance range **– The optimal impedance of the screen grid is approximately 18.5 % of the anode load impedance (The Ultra-linear configuration is also known as "Distributed Load operation").**

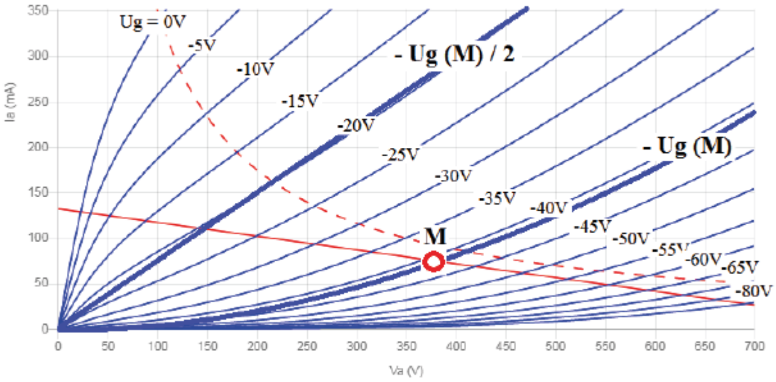


- If the impedance of the screen grid is optimal, the output power is almost equal to the maximum power delivered by the same tube in the pentode mode of operation.
- The internal impedance of the tube is low (almost equal to the internal resistance of the tube in triode mode).
- High level IM and THD distortion is minimal at high output power.
- Low level IM and THD distortion is low at low output power.



The optimal feed-back voltage of the screen grid is approximately 43% of the anode signal voltage. The tap of the primary of the output transformer connected to the screen grid is 43% of the total number of turns of the primary of the output transformer, counting from the end of the primary winding which is connected to the power supply or the center tap of the output transformer of the push-pull amplifier.

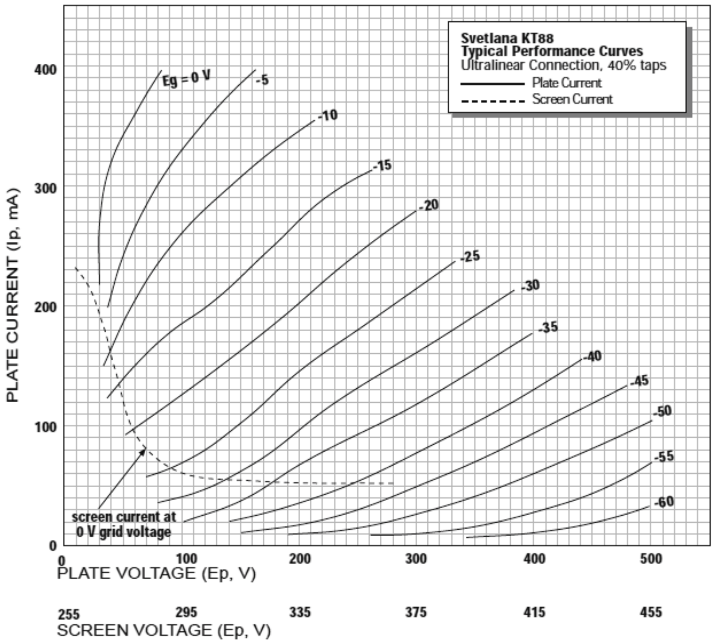
By analyzing the diagram of typical characteristics
I_a = f(Ua)│U_{g1}= ctc UL mode of the KT88 tetrode, it can be noted that one characteristic is **linear** and its bias voltage is **half** of the quiescent point M bias voltage. This is the optimal quiescent (working) point because it is located in the middle of the range of characteristics I_a = f(U_a).



Ultra linear tube configuration “provides the high power output of tetrodes with low internal impedance such as is normally obtained from triodes, while distortion figures are equal or better than the extremes of operations.” *Hafler and Keroes*
Ultra linear pentode (tetrode) configuration can be used in single ended and push-pull output amplifiers.
UL SE amplifier operates in class A.
UL PP amplifier can operate in class A or class AB (Hi End audio amplifier).

* Detailed UL mode of operation published by: LinearAudio **http://www.linearaudio.net** .
Author Mr. **Rudolf Moers**

Typical characteristic diagram $I_a = f(U_a)\ |U_{g1}$ = ctc. - UL mode of the beam tetrode



Example of Tube Manufacturer's Application Note - UL mode (Distributed Load):

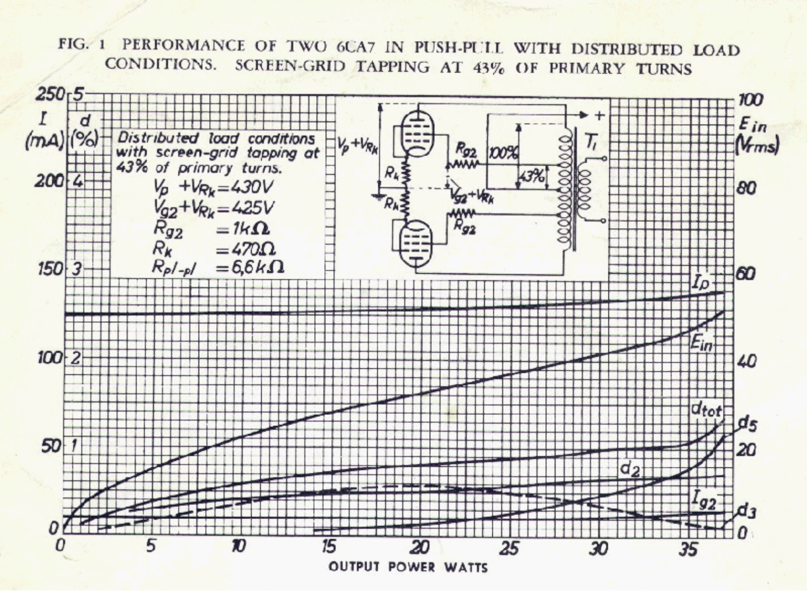


FIG. 1 PERFORMANCE OF TWO 6CA7 IN PUSH-PULL WITH DISTRIBUTED LOAD CONDITIONS. SCREEN-GRID TAPPING AT 43% OF PRIMARY TURNS

Example of Tube Manufacturer's Application Note - UL mode (Distributed Load):

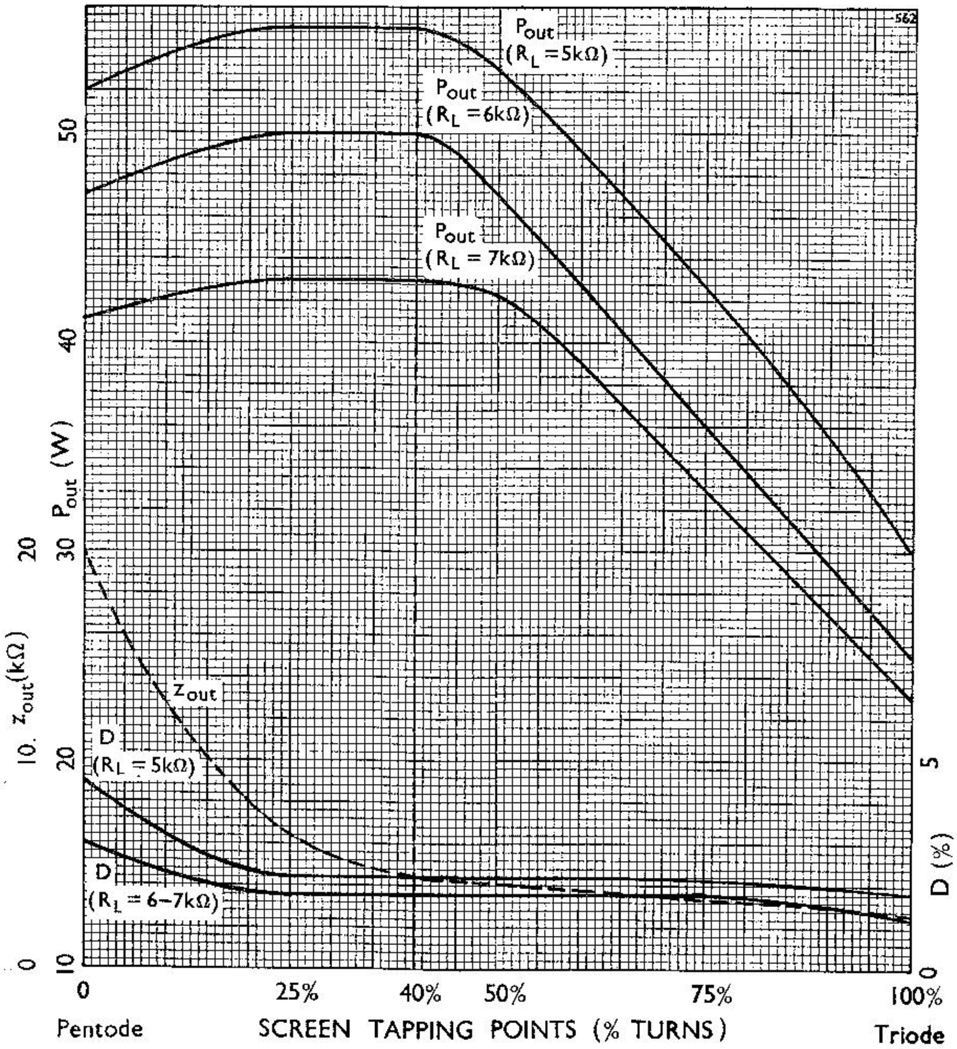
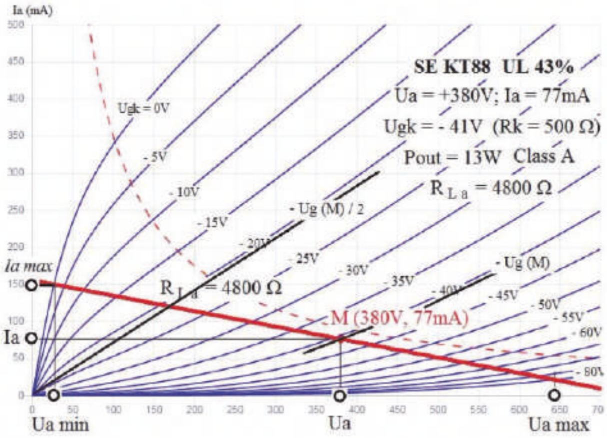
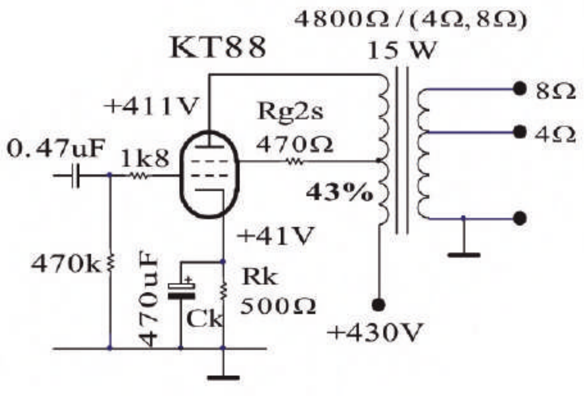


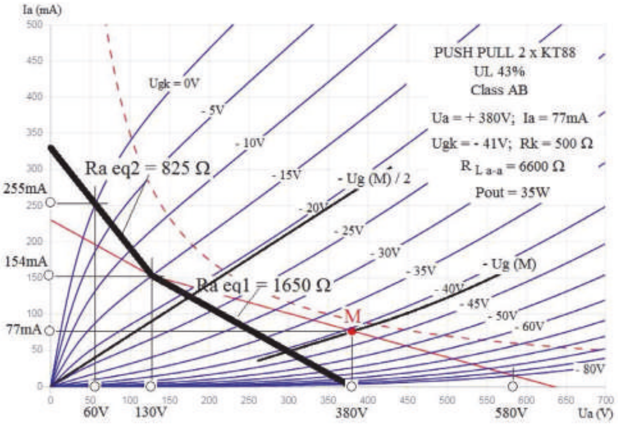
Fig. 1. Output power, distortion and output impedance of KT88 ultra-linear output stage at various positions of the screen taps from 0% to 100% of each half-primary from the centre tap.

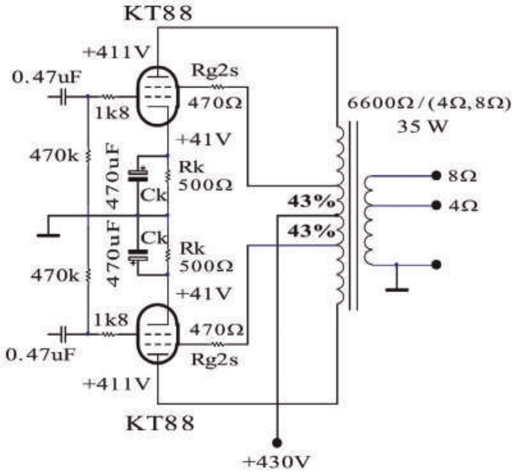
Example:
Single-Ended Class-A Ultra linear (43%) output amplifier



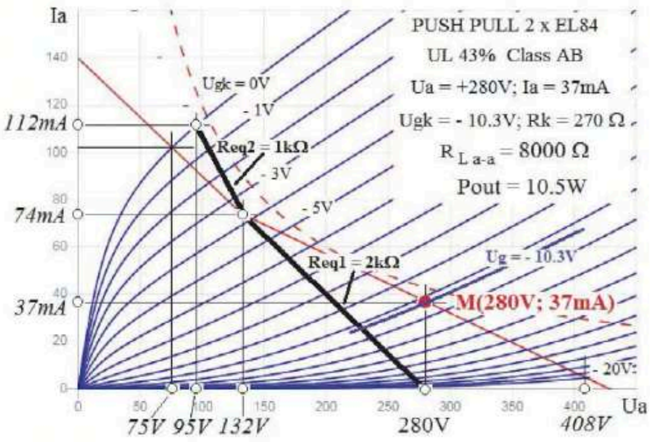


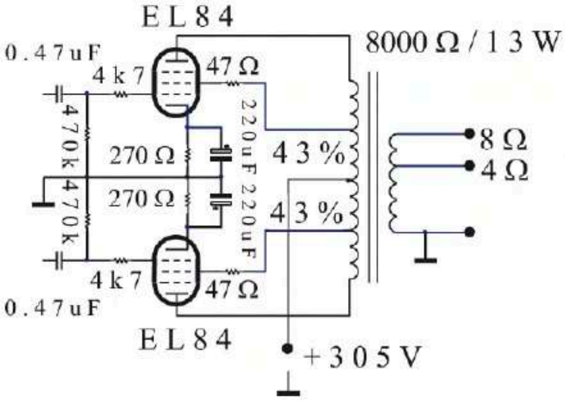
Example:
Push-Pull Class-AB Ultra linear (43%) output amplifier





Example:
Push Pull Class AB Ultra linear (43%) output amplifier





SUMMARY

There are a wide variety of amplifier stage configurations that can be applied in tube amplifiers. Some of the amplifier output stage configurations are not covered in the text above (they are too complicated for DIY building or do not fulfill the requirements of Hi End amplifiers).

Instead of a text summary, some notes on the application of some tubes are presented that can be helpful in the amplifier design process. The presented numerical values should not be accepted as absolutely correct and only possible, they can be useful as a starting point in the amplifier design process. Always consult the tube data sheets (but be careful using the manufacturer's examples of tube application notes as tube manufacturers sometimes publish tube application notes at the tube's maximum operating conditions).

Output stage circuit:

*Tube **300B**:*
SE *Ua = 350V; Ia = 80 mA; Ugk = - 71V (Rk = 880 Ω); $R_{L\ a}$ = 3000 Ω - 3500 Ω; Pout = 7.5W*
PP *Ua = 350V; Ia = 80 mA; Ugk = - 71V (Rk = 880 Ω); $R_{L\ a\text{-}a}$ = 5000 Ω - 5600 Ω; Pout = 17.5W*

*Tube **EL84**:*
PP UL (43%) Class A *Ua = 275V; Ia = 36 mA; Ugk = - 10V (Rk = 270 Ω); $R_{L\ a\text{-}a}$ = 13000 Ω; Pout = 8.5W*
PP UL 43%)(Class AB *Ua = 275V; Ia = 36 mA; Ugk = - 10V (Rk = 270 Ω); $R_{L\ a\text{-}a}$ = 8000 Ω; Pout = 10W*

Tube ***EL34****:*

SE *TRIODE Ua = 360V; Ia = 60 mA; Ugk = - 27.5V (Rk = 460 Ω); $R_{L\ a}$ = 4800 Ω; Pout = 6W*

SE UL (43%) *Ua = 360V; Ia = 60 mA; Ugk = - 27.5V (Rk = 460 Ω); $R_{L\ a}$ = 5000 Ω; Pout = 11W*

PP UL (43%) Class A *Ua = 360V; Ia = 60 mA; Ugk = - 27.5V (Rk = 460 Ω); $R_{L\ a\text{-}a}$ = 11000 Ω; Pout = 20W*

PP UL (43%) Class AB *Ua = 360V; Ia = 60 mA; Ugk = - 27.5V (Rk = 460 Ω); $R_{L\ a\text{-}a}$ = 6600 Ω; Pout = 28.5W*

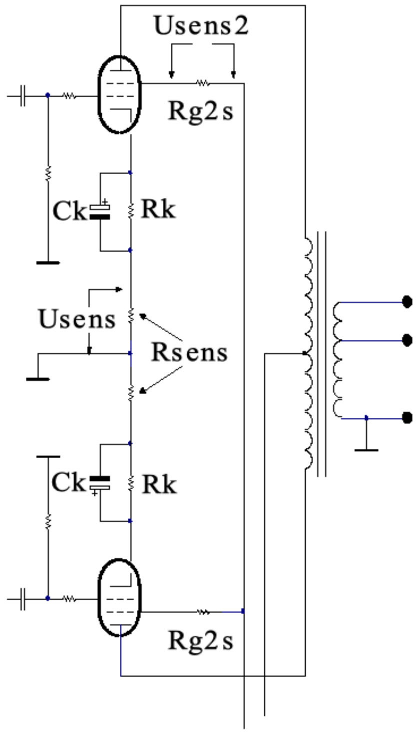
Tube ***KT88****:*

SE *TRIODE Ua = 390V; Ia = 75 mA; Ugk = - 42.5V (Rk = 560 Ω); $R_{L\ a}$ = 4500 Ω; Pout = 8W*

SE UL (43%) *Ua = 380V; Ia = 73 mA; Ugk = - 41.5V (Rk = 560 Ω); $R_{L\ a}$ = 4500 Ω; Pout = 13W*

PP UL (43%) Class A *Ua = 380V; Ia = 73 mA; Ugk = - 41.5V(Rk = 560 Ω); $R_{L\ a\text{-}a}$ = 9000 Ω; Pout = 27W*

PP UL (43%) Class AB *Ua = 380V; Ia = 73 mA; Ugk = - 41.5V (Rk = 560 Ω); $R_{L\ a\text{-}a}$ = 6600 Ω; Pout = 35W*



Useful to know

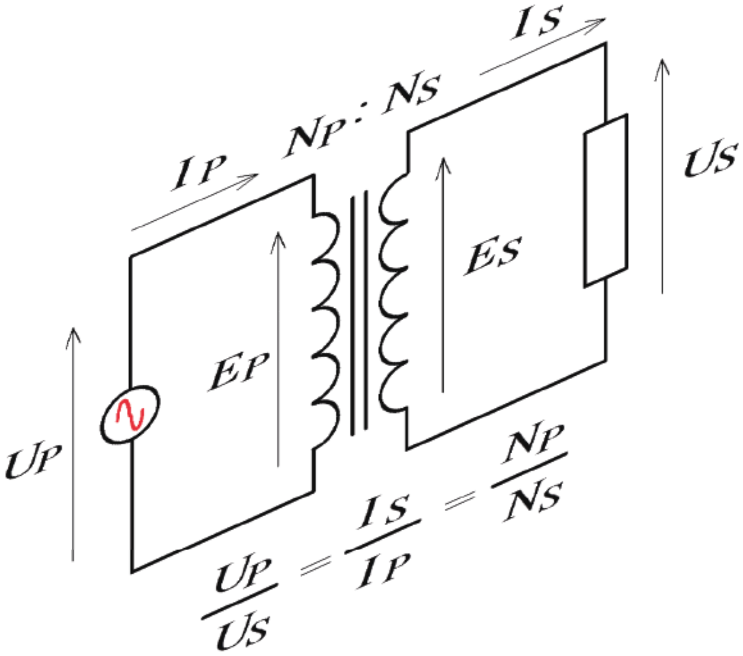
Connect low resistance resistors in the cathode circuits of the output tubes. By measuring the DC voltage across the resistors, the cathode currents (DC) of the output tubes can be calculated: $I_k = U_{sens} \ / \ R_{sens}$ (the cathode DC current is equal to the anode DC current of the triode or is equal to the sum of DC currents of the anode and the screen grid of the pentode or tetrode).

The measurement procedure must be performed under the condition that no driving AC signal(s) is (are) applied to the control grid of the output tube(s). Due to the low resistance of the resistor (of the ordered of 10 Ω) there are no consequences to the amplitude characteristic of the amplifier.

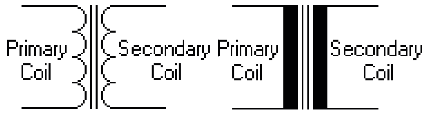
Also, the screen grid current (DC) can be calculated by measuring the voltage across the screen grid resistor.

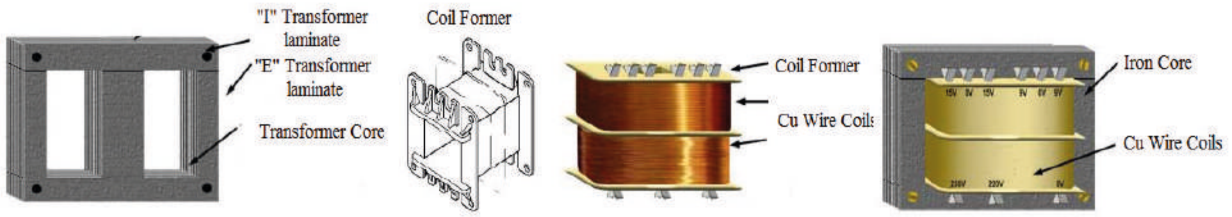
(the anode DC current can be calculated as the difference between the cathode current and the screen grid current).

Chapter 5 • Transformers

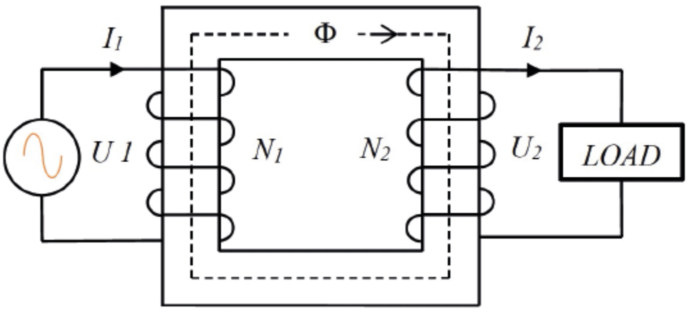


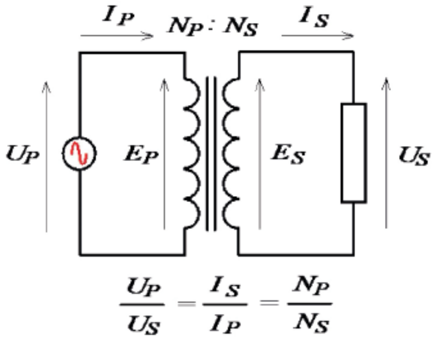
5.0 TRANSFORMERS





5.1 TRANSFORMERS – BASIC (ELEMENTARY)





A transformer is a device in which two circuits are coupled by a magnetic field that is linked to both. There is no conductive connection between the circuits.

In general, transformer is electronic component that consists of the two or more coils wound on the same magnetic core. Also, a transformer is an electrical device that transfers electrical energy between two or more circuits through electromagnetic induction.

When a signal of a defined form, frequency, voltage and power from a generator is connected to the Primary coil, electric current passes through the Primary coil. A signal of the same shape and frequency appears at the secondary side of the transformer. But the voltage at the secondary side and the current passing through the Secondary coil and the load connected to the Secondary need not be the same as on the primary side. The Secondary voltage and current (also the impedance of the Primary and Secondary) depend on the turns ratio of the Primary and Secondary coils, but the powers of the primary and secondary sides are the same.

When electric current flows through the coil it generates a magnetic field.
The magnetic field strength *or* ***magnetizing force*** *(also known as the magnetic potential gradient or magnet field intensity) is defined as the* ***magnetomotive force per unit length of the magnetic circuit*** *produced by the current flowing in the coil and it is denoted as* ***H:***

H = magnetomotive force (mmf) per unit length of the magnetic circuit.

mmf *is the force produced by the current* ***I*** *flowing through the* ***N*** *turns of the coil.*
If the length of the magnetic circuit is denoted by **MPL,** the magnetic field strength or magnetizing force H is:

$$H = \frac{4 \times \pi \times I \times N}{10 \times MPL} = \frac{1.256 \times I \times N}{MPL} \; (Oersted)$$

(#Note: Unit Oersted is equal to 1.257 ampère - turns per cm)
The result of the action of this magnetic field is the magnetic flux that is produced inside the magnetic material. The intensity of this flux is called ***flux density (B)****.*

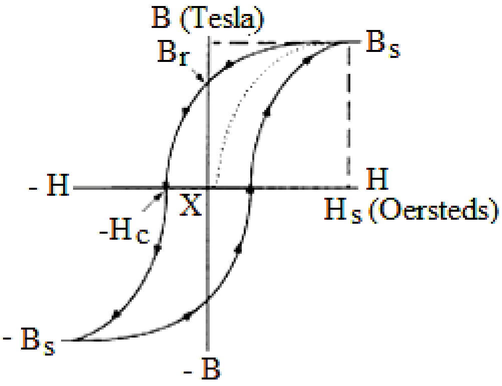
The total flux is:

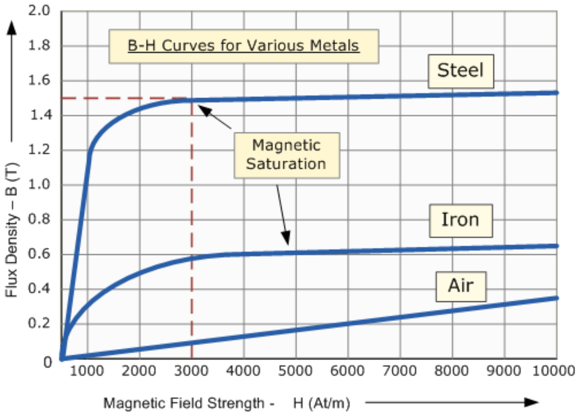
$$\emptyset = B \times S = \mu \times H \times S = \frac{4 \times \pi \times N \times I \times \mu \times S}{10 \times MPL}$$

B – *magnetic flux density, Induction*
Units: *1 T = 10^4 Gs (T = Tesla, Gs = Gauss)*

B - H relation

A graphical representation of the dependence of the produced magnetic flux on the magnetizing force for each magnetic material is known as the ***B-H curve*** *of the material.*





$$B = \mu \times H = \mu_r \times \mu_0 \times H$$

As illustration only

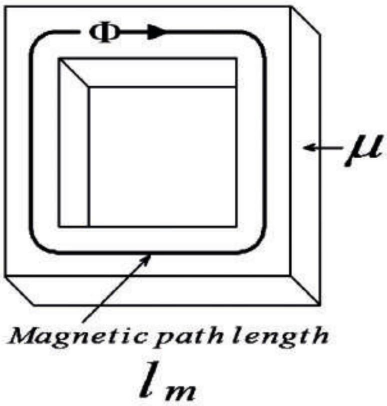
μ_0 – *permeability of the vacuum ($4x\pi x10^{-7}$ H/m)*
The permeability of magnetic materials is usually specified as relative permeability μ_r to the permeability of free space (denoted by 1).
The permeability of a magnetic material is the ability of the material to increase the flux intensity or flux density within the material when electric current flows through a coil wound around the magnetic material and produces a magnetizing force.

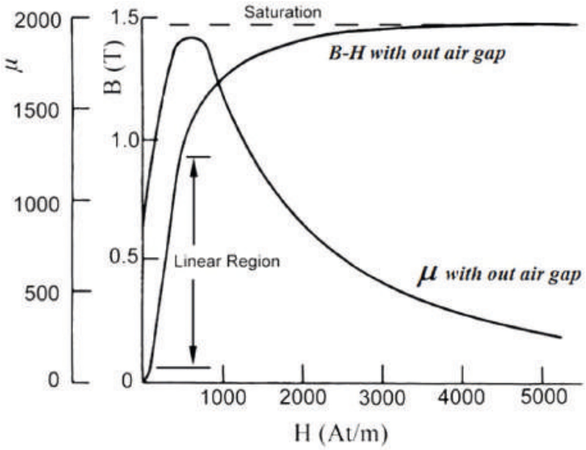
The ***inductance*** *of the coil of N turns wound on a core of cross-section S and permeability μ is:*

$$L = \frac{4 \times \pi \times N^2 \times \mu \times S[cm^2]}{MPL[cm]} \times 10^{-8} = \frac{1.256 \times N^2 \times \mu \times S[cm^2]}{MPL[cm]} \times 10^{-8}\ Henry[H]$$

$$\mu_{eff} \approx \mu$$

B – H curve of a **closed** core of a magnetic material with magnet permeability μ and a continuous closed magnetic path.



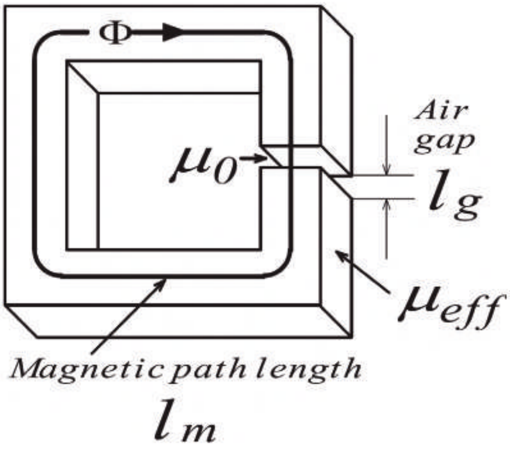


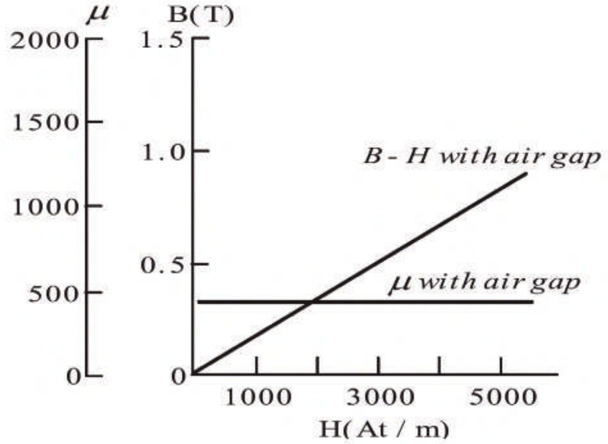
The magnetic force H produced by the current flowing through the coil generates a magnetic flux B and magnetizes the transformer core. At a low magnetic force H produced by the current flowing through the coil, the B-H curve is curved. As H increases, the B-H curve becomes very linear up to a value when it curves again, and at very high H it becomes flat – the magnetic core is ***saturated****. When the magnetic core is saturated, the permeability is reduced to zero value (μ= 0), and thus the inductivity of the coil is reduced to zero value (L=0).*
Once this happens, any further increase in H, will not increase the flux. This situation causes the loss of the basic function of the transformer. *A transformer is normally designed to ensure that the magnetic flux density is below the level that would cause saturation.*

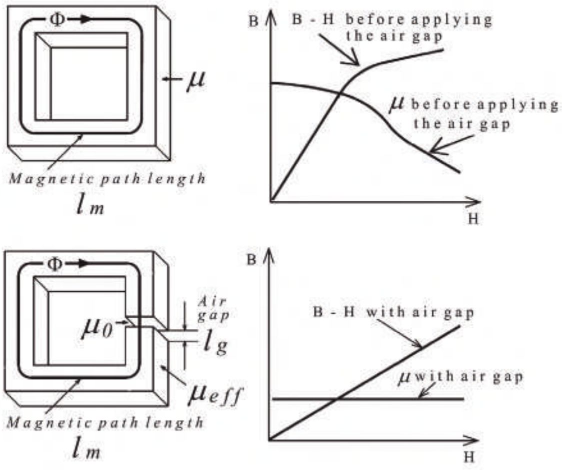
When an ***air gap*** *is placed in the magnetic path, the B-H curve changes. It becomes more linear in a wide H range and core saturation is avoided. The effective permeability is almost independent of H and is constant in most of the B-H diagram.*
Effective permeability of an air- gaped magnetic circuit:

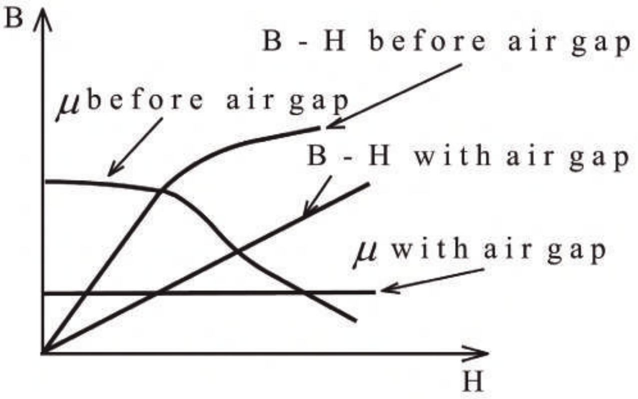
$$\mu_{eff} = \frac{\ell_m}{\frac{\ell_m}{\mu} + \frac{\ell_g}{\mu_g}} = \frac{1}{\frac{\ell_g}{\ell_m} + \frac{1}{\mu}}$$

* $\mu_g = \mu_0 = 1$ *(permeability of air gap is equal to vacuum permeability)*
Effective permeability is lower than the permeability of the transformer core laminate material: $\mu_{eff} < \mu$.
The most important effect of the air gap is to prevent saturation of the transformer core caused by high DC current or large AC signal.









Electric Circuit		Magnetic Circuit	
Quantity	Unit	Quantity	Unit
E.m.f	Volt	m.m.f	Ampere
		Magnetic field strength H	Oersted
Current	Ampere	Magnetic flux Φ	Maxwell
Current density	Ampere/m^2	Magnetic flux density **B**	Gauss
Resistance	Ohm	Reluctance **S**	Ampere/Maxwell
Current = e.m.f./ resistance		Flux = m.m.f./reluctance	

Unit	Symbol	Definition	To convert to CGS	Multiply by
Henry (H)	**L**	The *inductance* of a closed circuit in which an e.m.f. of 1 V is produced when the electric current varies at a rate of 1 A/s.		
Weber (Wb)	Φ	The *magnetic flux* which, when linking a circuit of one turn, produces in it an e.m.f. of 1 V when it is reduced to zero at a uniform rate in 1 s.	Maxwell & lines	1×10^8
Ampere/meter	**H**	ampère-turns per metre. (ampère-turns per cm).	Oersted	0.01257 (1.257)
Tesla (T)	**B**	The magnetic flux density equal to 1 Wb/m^2 .	Gauss & Lines/cm2	10^4
Henry/meter	μ_0	Permeability of free space $4\pi \times 10^{-7}$.	Gauss / Oerstead	795774.72 (= unity)

Transformer equations

If an alternating current source (AC) is connected to the Primary of the transformer, electromotive forces **(emf**) (caused by magnetic flux) are induced in the Primary and Secondary coils:

$$emf_p = 4.44 \times 10^{-4} \times f \times N_p \times B_m \times S_{eff} \tag{501.1}$$

$$emf_s = 4.44 \times 10^{-4} \times f \times N_s \times B_m \times S_{eff} \tag{501.2}$$

f – frequency [Hz]
N_p – Number of turns of the Primary
N_s – Number of turns of the Secondary
B_m – Flux Density (Magnetic Flux) [T]
S_{eff} – Cross-sectional area of the Transformer Core [cm^2]

The electromotive forces (*emf) of the primary and secondary differ from the supply voltage U_p of the primary and the induced voltage U_s at the secondary for voltage drops on the coils, so it must be taken into account emf = 0.9 x U :*

$U_P \approx emf_P$ and $U_S \approx emf_S$:

$$\frac{U_s}{U_p} = \frac{N_s}{N_p} = n \qquad (501.3)$$

n – turns ratio
N_p – number of turns of the Primary
N_s – number of turns of the Secondary

If the secondary is loaded, current will flow through the Primary and Secondary coils:

$$\frac{I_s}{I_p} = \frac{N_p}{N_s} = \frac{1}{n} \qquad (501.4)$$

The inductance of the coil (L)
The inductance of a coil of **N** turns wound on a magnetic core (cross-section $\mathbf{S_{eff}}$ and magnetic path length **MPL**) is:

$$\boldsymbol{L} = 1.257 \times 10^{-8} \times \mu \times N^2 \times \frac{S_{eff}}{MPL} \ (H) \qquad (501.5)$$

Unit: H = (Henry)

OTHER RELATIONS

$$\frac{U_p}{U_s} = \frac{I_s}{I_p} = \frac{N_p}{N_s} = \sqrt{\frac{L_p}{L_s}} = \sqrt{\frac{Z_p}{Z_s}} = \mathrm{n} \qquad (501.6)$$

L_p , L_s – Inductance of the Primary and Secondary
Z_p , Z_s – Impedance of the Primary and Secondary
#Note: Powers at the input (Primary) and output (Secondary) of the transformer are equal.

5.2 MAGNETIC CORE

Magnetic cores, in common application, are made from ferroalloy laminates of a standardized (Table 501.1) shape. A common shape is **E / I** laminate:

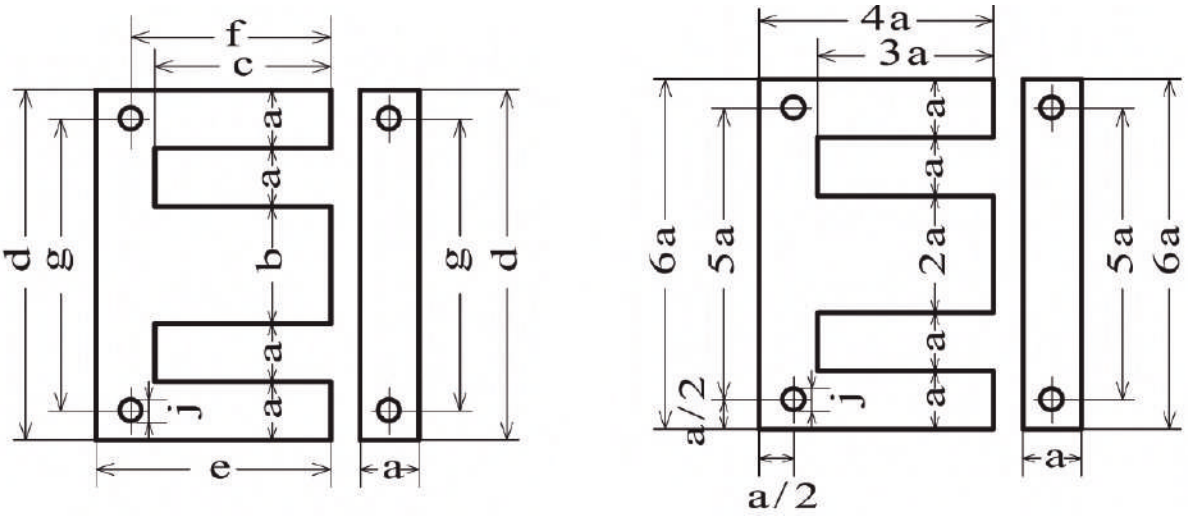


Fig. 5-01

E I Type:
Standard type, dimension ratios:
b = 2 x a
c = 3 x a
e = 4 x a
d = 6 x a

Each piece of the laminates is insulated.
Laminate thickness: **0.1 mm, 0.35 mm, 0.5 mm**.

If the laminates are grouped together they form a transformer core.

They can be stacked in two ways:

- All **E** laminates stack in the same direction and all **I** laminates stack in a same direction. Fig. 5-02. Usually, insulation – a non magnetic material or so called ” air gap”, is inserted between **E** and **I** stacks.
- E laminates are stacked one by one in the opposite direction. **I** laminates are inserted between **E** laminates Fig. 5-02.

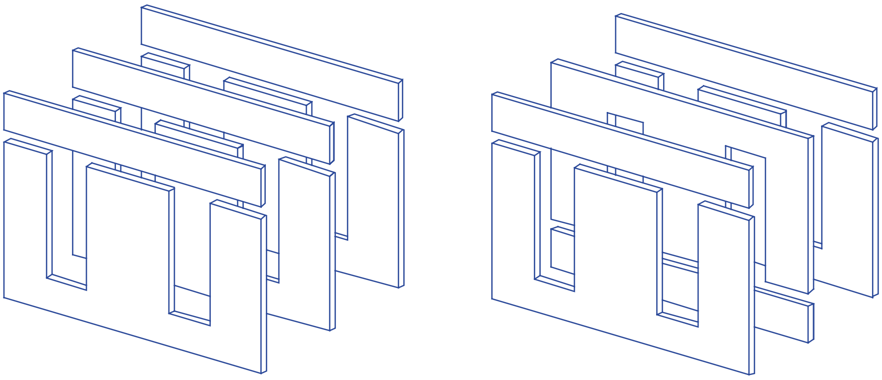


Fig. 5-02

The coil former on which the coils of Cu wire are wound must be made of mechanically strong and electrically insulating material.

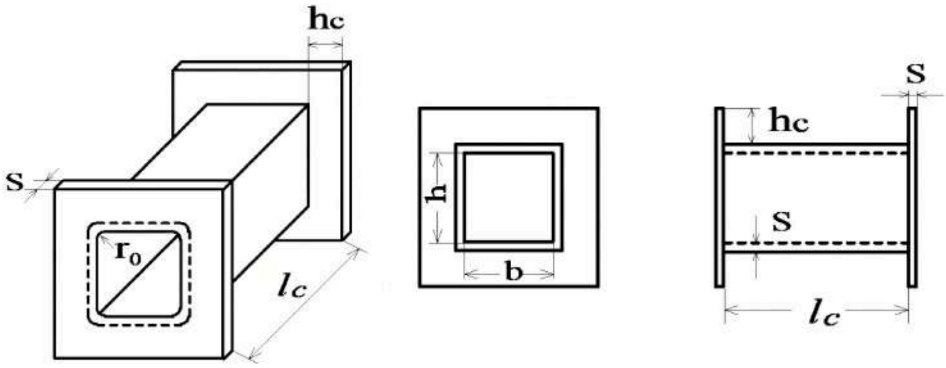
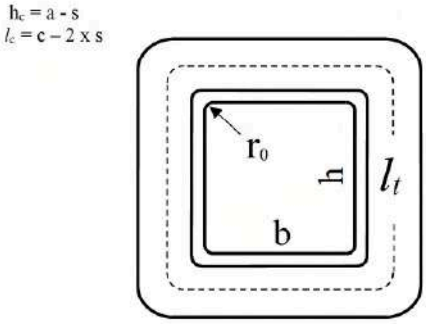
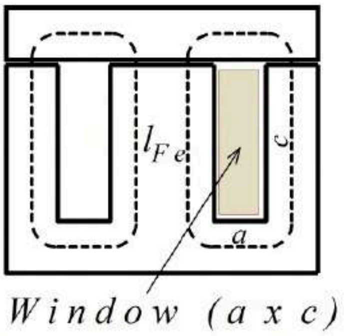
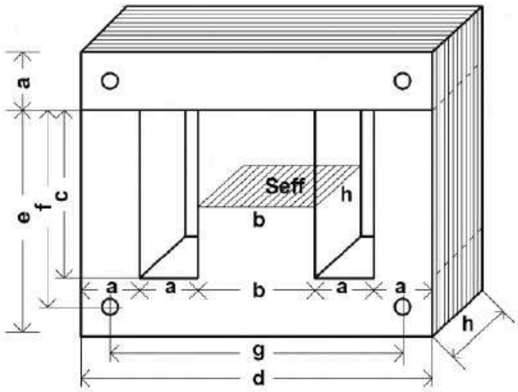


Fig. 5-03







5.3 Some general parameters used to design transformers:

Effective cross-section of the core **S_{eff} :**

$$\boldsymbol{S_{eff}}(cm^2) = \boldsymbol{k_i} \times \boldsymbol{b}(cm) \times \boldsymbol{h}(cm) \qquad (501.7)$$

h – stack height [cm]
k_i – stacking coefficient (depends on the thickness of laminates and insulating material):

$0.85 \le k_i \le 0.98$; ($k_i = 0.96$)

Cross-section of the core window **S_w :**

$$\boldsymbol{S_w}(cm^2) = \boldsymbol{c}(cm) \times \boldsymbol{a}(cm) \qquad (501.8)$$

Magnetic circuit length **MPL = ℓ_{Fe}:**

$$\boldsymbol{\ell_{Fe}}(cm) = \boldsymbol{2} \times [\boldsymbol{c}(cm) + \boldsymbol{3} \times \boldsymbol{a}(cm)]$$

$$\boldsymbol{\ell_{Fe}}(\boldsymbol{cm}) = 2 \times \boldsymbol{c}(cm) + 2 \times \boldsymbol{a}(cm) + \boldsymbol{\pi} \times \boldsymbol{a}(cm)\ (\boldsymbol{more\ accurate})\ \text{or:}$$

$$\boldsymbol{\ell_{Fe}}(\boldsymbol{cm}) = 4 \times b(cm) + \frac{\pi \times b}{2}$$

A simplified equation used in practice:

$$\boldsymbol{\ell_{Fe}}(cm) = \boldsymbol{5.57} \times \boldsymbol{b}(cm) \qquad (501.9)$$

Volume of the core **V_{Fe}:**

$$\boldsymbol{V_{Fe}}(cm^3) = \boldsymbol{S_{eff}}(cm^2) \times \boldsymbol{\ell_{Fe}}(cm) \qquad (501.10)$$

Mass of the core (weight) **G_{Fe}** :

$$G_{Fe}(gr) = V_{Fe}(cm^3) \times D_{Fe}\left(\frac{g}{cm^3}\right) = S_{eff}(cm^2) \times \ell_{Fe}(cm) \times D_{Fe}\left(\frac{g}{cm^3}\right) \quad (501.11)$$

D_{Fe} – density of ferromagnetic material
Commonly used material: **4% Si steel.**
Density: **$D_{Fe(4\%Si)} = 7.65$ g / cm^3**

If the core dimensions, the coil former dimensions, the diameters of the insulated Cu wire and the numbers of turns are known, the calculation proceeds as follows:

Length of the coil:

$$\ell_c = c - (4 \times s) \quad (501.12)$$

s – thickness of the coil former walls **[(1 ÷ 2) mm]**

Number of turns per layer:

$$N_{lc} = \frac{l_c}{d_w \times k} \quad (501.13)$$

d_w – outside (overall) wire diameter
k - coefficient (depends on the diameter of wire (usually **k = 1.03**)

Number of layers:

$$z = \frac{N}{N_{lc}} \quad (501.14)$$

N = Total number of turns

Z is the first higher integer.

Radial dimension of the each coil (coil height):

$$h_n = z \times d_w + (z-1) \times \Delta \quad (501.15)$$

Δ – insulation thickness between layers **(0.05mm).**

Total radial dimension (height) of all coils:

$$h_0 = \sum h_n + \sum \Delta_{j(p-s)} + s + \Delta_z + \Delta_0 \quad (501.16)$$

Δ_z **–** gap between the coil former and the magnetic core
Σh_n **–** the sum of the radial dimensions of the coils
$\Sigma\Delta_{j(p-s)}$ **–** the sum of the thickness of the insulation between the coils
Δ_0 **–** thickness of outside or final insulation

 # Condition: $h_0 < hc = a - s$ must be fulfilled.

DC resistance of the coil:

$$R_c = \rho \times \frac{4 \times l_w}{\pi \times d^2} \quad (501.17)$$

d – wire diameter (mm)

ρ – specific resistance of the wire **(For Cu wire** $\rho = 0.0175\Omega\frac{mm^2}{m}$ **)**

l_w – length of the coil wire (m):

$$l_w = N \times l_t \qquad (501.18)$$

N – number of turns**.**

l_t**– mean** length per turn (MLT):

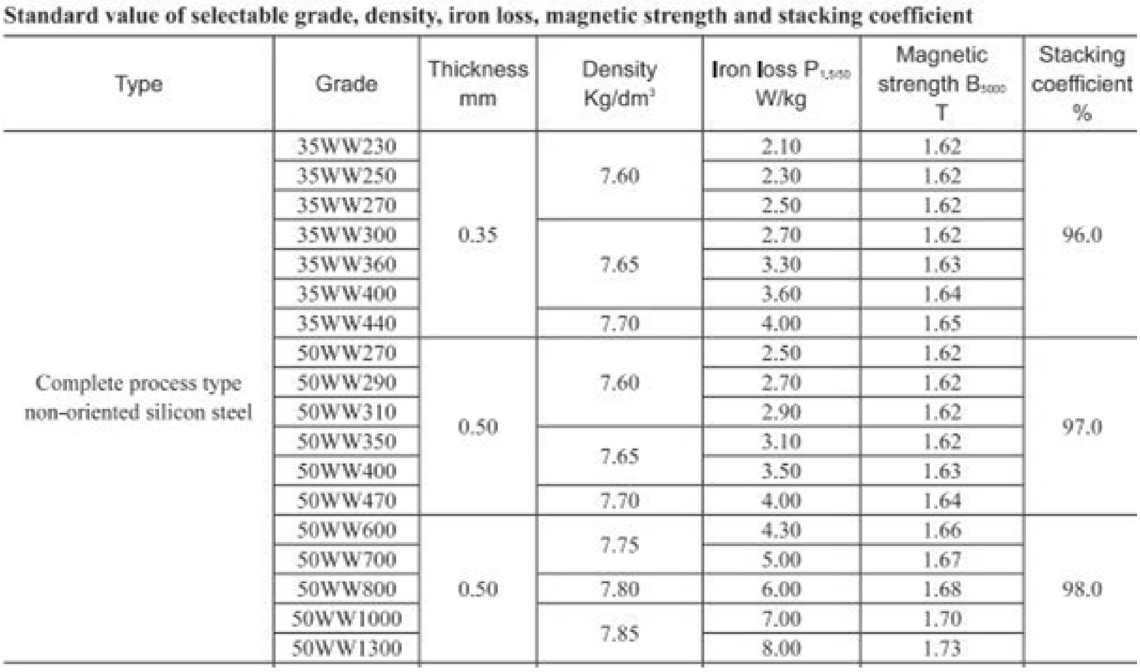
$$l_t = (2 \times b) + (2 \times h) + (\pi \times h_0) + (4 \times s)$$

$$l_t = 2 \times (b + h) + 2 \times \pi \times (r_0 + \frac{h_0}{2}) \qquad (501.19)$$

where $\mathbf{r_0}$ **–** radius of the coil former (1 mm)

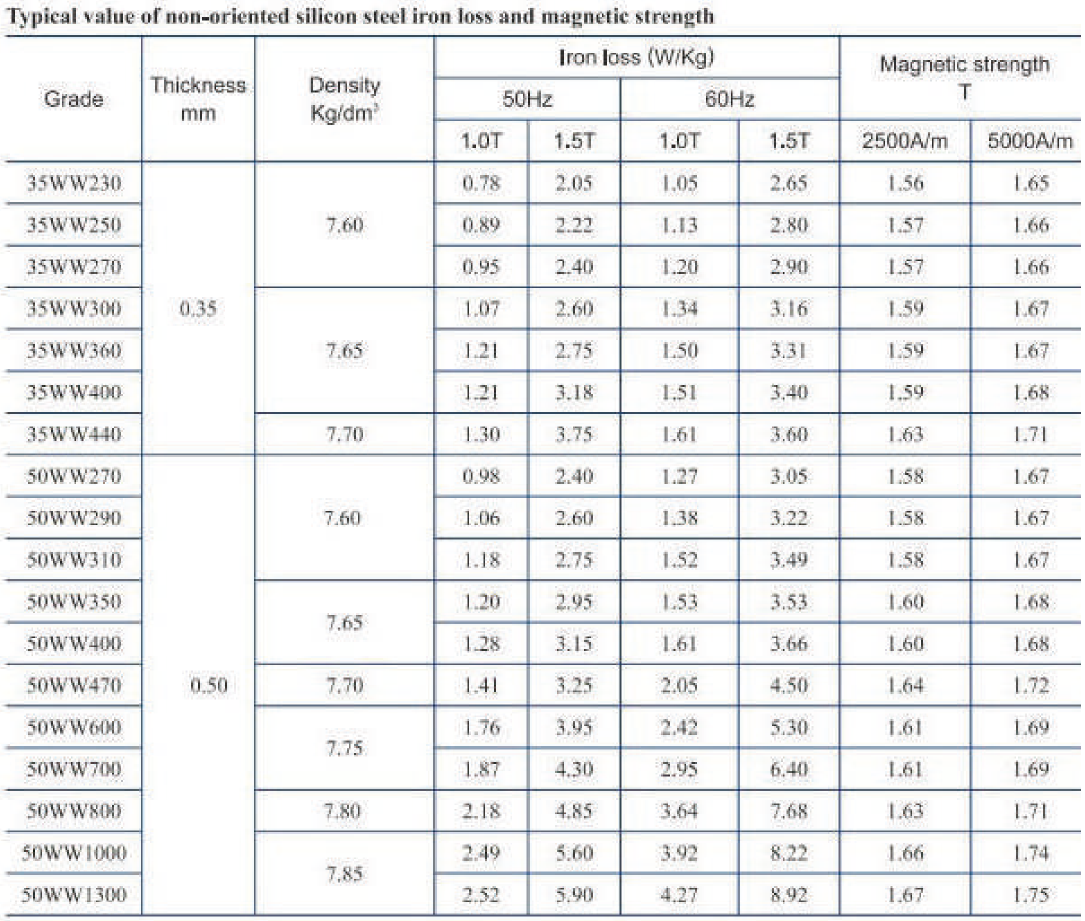
Table 501.1 **E I laminate.**

Type	**IEC**	**a [mm]**	**b [mm]**	**c [mm]**	**d [mm]**	**e [mm]**	**j [mm]**	**Magnetic circuit length ℓ_{Fe} [cm]**	**Window cross section Sw [cm²]**
EI 30	YEI 1-10	5.0	10.0	15.0	30.0	20.0	0.0	**5.57**	**0.75**
EI 38	YEI 1-13	6.4	12.8	19.2	38.4	25.6	0.0	**7.13**	**1.2288**
EI 42	YEI 1-14	7.0	14.0	21.0	42.0	28.0	3.5	**7.8**	**1.47**
EI 48	YEI 1-16	8.0	16.0	24.0	48.0	32.0	3.5	**8.9**	**1.92**
EI 54	YEI 1-18	9.0	18.0	27.0	54.0	36.0	3.5	**10.02**	**2.43**
EI 60	YEI 1-20	10.0	20.0	30.0	60.0	40.0	3.5	**11.14**	**3.00**
EI 66	YEI 1-22	11.0	22.0	33.0	66.0	44.0	4.5	**12.25**	**3.63**
EI 75	YEI 1-25	12.5	25.0	37.5	75.0	50.0	4.5	**13.92**	**4.6875**
EI 78		13.0	26.0	39.0	78.0	52.0	4.5	**14.48**	**5.07**
EI 84	YEI 1-28	14.0	28.0	42.0	84.0	56.0	4.5	**15.6**	**5.88**
EI 90		15.0	30.0	45.0	90.0	60.0	4.5	**16.71**	**6.75**
EI 96	YEI 1-32	16.0	32.0	48.0	96.0	64.0	5.5	**17.82**	**7.68**
EI 105		17.5	35.0	52.5	105.0	70.0	5.5	**19.5**	**9.1875**
EI 108	YEI 1-36	18.0	36.0	54.0	108.0	72.0	5.5	**20.05**	**9.72**
EI 120	YEI 1-40	20.0	40.0	60.0	120.0	80.0	7.0	**22.28**	**12.00**
EI 135		22.5	45.0	67.5	135.0	90.0	7.5	**25.06**	**15.1875**
EI 150N	YEI 1-50	25.0	50.0	75.0	150.0	100.0	8.0	**27.85**	**18.75**
EI 174	YEI 1-58	29.0	58.0	87.0	174.0	116.0	10.5	**32.3**	**25.23**
EI 180		30.0	60.0	90.0	180.0	120.0	10.0	**33.42**	**27.00**
EI 192	YEI 1-64	32.0	64.0	96.0	192.0	128.0	11.0	**35.65**	**30.72**
EI 216	YEI 1-72	36.0	72.0	108.0	216.0	144.0	11.5	**40.1**	**38.88**
EI 240	YEI 1-80	40.0	80.0	120.0	240.0	160.0	11.5	**44.56**	**48.00**



Standard value of selectable grade, density, iron loss, magnetic strength and stacking coefficient

Type	Grade	Thickness mm	Density Kg/dm³	Iron loss $P_{1.5/50}$ W/kg	Magnetic strength B_{5000} T	Stacking coefficient %
Complete process type non-oriented silicon steel	35WW230	0.35	7.60	2.10	1.62	96.0
	35WW250			2.30	1.62	
	35WW270			2.50	1.62	
	35WW300		7.65	2.70	1.62	
	35WW360			3.30	1.63	
	35WW400			3.60	1.64	
	35WW440		7.70	4.00	1.65	
	50WW270	0.50	7.60	2.50	1.62	97.0
	50WW290			2.70	1.62	
	50WW310			2.90	1.62	
	50WW350		7.65	3.10	1.62	
	50WW400			3.50	1.63	
	50WW470		7.70	4.00	1.64	
	50WW600	0.50	7.75	4.30	1.66	98.0
	50WW700			5.00	1.67	
	50WW800		7.80	6.00	1.68	
	50WW1000		7.85	7.00	1.70	
	50WW1300			8.00	1.73	



Typical value of non-oriented silicon steel iron loss and magnetic strength

Grade	Thickness mm	Density Kg/dm³	Iron loss (W/Kg)				Magnetic strength T	
			50Hz		60Hz			
			1.0T	1.5T	1.0T	1.5T	2500A/m	5000A/m
35WW230	0.35	7.60	0.78	2.05	1.05	2.65	1.56	1.65
35WW250			0.89	2.22	1.13	2.80	1.57	1.66
35WW270			0.95	2.40	1.20	2.90	1.57	1.66
35WW300		7.65	1.07	2.60	1.34	3.16	1.59	1.67
35WW360			1.21	2.75	1.50	3.31	1.59	1.67
35WW400			1.21	3.18	1.51	3.40	1.59	1.68
35WW440		7.70	1.30	3.75	1.61	3.60	1.63	1.71
50WW270	0.50	7.60	0.98	2.40	1.27	3.05	1.58	1.67
50WW290			1.06	2.60	1.38	3.22	1.58	1.67
50WW310			1.18	2.75	1.52	3.49	1.58	1.67
50WW350		7.65	1.20	2.95	1.53	3.53	1.60	1.68
50WW400			1.28	3.15	1.61	3.66	1.60	1.68
50WW470		7.70	1.41	3.25	2.05	4.50	1.64	1.72
50WW600		7.75	1.76	3.95	2.42	5.30	1.61	1.69
50WW700			1.87	4.30	2.95	6.40	1.61	1.69
50WW800		7.80	2.18	4.85	3.64	7.68	1.63	1.71
50WW1000		7.85	2.49	5.60	3.92	8.22	1.66	1.74
50WW1300			2.52	5.90	4.27	8.92	1.67	1.75

Material	Permeability μ_{max}	Saturation limit (Bmax) Gauss
4% Si steel non – oriented **NOSS** 0.35mm	8000 **(3000)**	18000 (13000) **(12000)**
4% Si steel grain – oriented **GOSS** 0.35mm	10000 – 50000 **(5000)**	5000 - 20000

$B_{ac} = B_{dc} = ½ B_{max}$
B_{max} = 1.5 T GOSS
B_{max} = 1.3 T NOSS
B_{max} = 1.0 T other

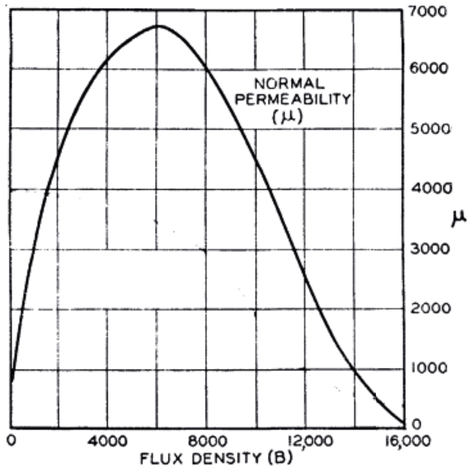


Table 501.2
Cooper wire standards:

No	**SWG** Ø **[mm]**	**area [mm²]**	Ω / 100 m	**AWG** Ø **[mm]**	**area [mm²]**	**RESISTANCE** Ω / 100m
10	3.25	8.30	0.211	2.59	5.26	0.332
11	2.95	6.84	0.256	2.31	4.17	0.418
12	2.64	5.47	0.32	2.05	3.30	0.530
13	2.34	4.30	0.407	1.83	2.63	0.665
14	2.03	3.24	0.541	1.63	2.08	0.839
15	1.83	2.63	0.665	1.45	1.65	1.060
16	1.63	2.08	0.839	1.29	1.31	1.339
17	1.42	1.58	1.105	1.15	1.04	1.685
18	1.22	1.17	1.497	1.02	0.82	2.142
19	1.02	0.82	2.142	0.91	0.65	2.691
20	0.91	0.65	2.691	0.81	0.52	3.396
21	0.81	0.52	3.396	0.72	0.41	4.298
22	0.71	0.40	4.420	0.64	0.32	5.440
23	0.61	0.29	5.988	0.57	0.26	6.858
24	0.56	0.25	7.105	0.51	0.20	8.567
25	0.51	0.20	8.567	0.45	0.16	11.00
26	0.46	0.16	10.53	0.40	0.13	13.93
27	0.42	0.14	12.63	0.36	0.10	17.19
28	0.38	0.11	15.43	0.32	0.08	21.76

29	0.35	0.10	18.19	0.29	0.07	26.50
30	0.32	0.08	21.76	0.25	0.05	35.65
31	0.29	0.07	26.49	0.23	0.04	42.12
32	0.27	0.06	30.56	0.20	0.03	55.70
33	0.25	0.05	35.65	0.18	0.025	68.77
34	0.23	0.04	42.12	0.16	0.020	87.04
35	0.21	0.036	50.53	0.14	0.015	113.7
36	0.19	0.029	61.72	0.13	0.013	131.8
37	0.17	0.023	77.10	0.11	0.0095	184.1
38	0.15	0.018	99.03	0.10	0.0079	222.8
39	0.13	0.013	131.8	0.09	0.0064	275.1
40	0.12	0.012	154.7	0.08	0.0050	348.2
41	0.11	0.0095	184.1	-	-	-
42	0.10	0.0079	222.8	-	-	-

SWG (Imperial Standard Wire Gauge)

AWG (American Wire Gauge)

According to IEC standard 60317 – 0 - 1

		Grade 1		Grade 2		Grade 3		Length (m/kg)			Area mm^2	Resistance 20°C Ω/m
Nominal conductor diameter	Conductor tolerance +/-	min increase	max overall diameter	min increase	max overall diameter	min increase	max overall diameter	Grade 1	Grade 2	Grade 3	Nominal	Nominal
0.090	0.003	0.008	0.105	0.015	0.113	0.022	0.120	16628	16121	15675	0.00636	2.6870
0.100	0.003	0.008	0.117	0.016	0.125	0.023	0.132	13453	13084	12759	0.00785	2.1765
0.112	0.003	0.009	0.130	0.017	0.139	0.026	0.147	10759	10463	10199	0.00985	1.7351
0.125	0.003	0.010	0.144	0.019	0.154	0.028	0.163	8663	8427	8213	0.01227	1.3929
0.140	0.003	0.011	0.160	0.021	0.171	0.030	0.181	6927	6743	6574	0.01539	1.1104
0.150	0.003	0.012	0.171	0.023	0.182	0.033	0.193	6040	5890	5739	0.01767	0.9673
0.160	0.003	0.012	0.182	0.023	0.194	0.033	0.205	5313	5179	5054	0.02011	0.8502
0.180	0.003	0.013	0.204	0.025	0.217	0.036	0.229	4204	4102	4006	0.02545	0.6718
0.200	0.003	0.014	0.226	0.027	0.239	0.039	0.252	3409	3335	3259	0.03142	0.5441
0.212	0.003	0.015	0.240	0.029	0.254	0.043	0.268	3032	2965	2897	0.03530	0.4843
0.224	0.003	0.015	0.252	0.029	0.266	0.043	0.280	2722	2665	2608	0.03941	0.4338
0.236	0.004	0.017	0.267	0.032	0.283	0.048	0.298	2447	2391	2339	0.04374	0.3908
0.250	0.004	0.017	0.281	0.032	0.297	0.048	0.312	2186	2139	2095	0.04909	0.3482
0.265	0.004	0.018	0.297	0.033	0.314	0.050	0.330	1948	1906	1866	0.05515	0.3099
0.280	0.004	0.018	0.312	0.033	0.329	0.050	0.345	1748	1713	1679	0.06158	0.2776
0.300	0.004	0.019	0.334	0.035	0.352	0.053	0.36	1524	1493	1479	0.07069	0.2418
0.315	0.004	0.019	0.349	0.035	0.367	0.053	0.384	1384	1358	1333	0.07793	0.2193
0.335	0.004	0.020	0.372	0.038	0.391	0.057	0.408	1223	1200	1179	0.08814	0.1939
0.355	0.004	0.020	0.392	0.038	0.411	0.057	0.428	1091	1072	1054	0.09898	0.1727
0.375	0.005	0.021	0.414	0.040	0.434	0.060	0.453	978	961	944	0.1104	0.1548
0.400	0.005	0.021	0.439	0.040	0.459	0.060	0.478	861	847	834	0.1257	0.1360
0.425	0.005	0.022	0.466	0.042	0.488	0.064	0.508	763	750	738	0.1419	0.1205
0.450	0.005	0.022	0.491	0.042	0.513	0.064	0.533	682	671	661	0.1590	0.1075
0.475	0.005	0.024	0.519	0.045	0.541	0.067	0.562	612	603	594	0.1772	0.09646
0.500	0.005	0.024	0.544	0.045	0.566	0.067	0.587	553	545	537	0.1963	0.08706
0.530	0.006	0.025	0.576	0.047	0.600	0.071	0.623	492	485	478	0.2206	0.07748
0.560	0.006	0.025	0.606	0.047	0.630	0.071	0.653	442	436	430	0.2463	0.06940
0.600	0.006	0.027	0.649	0.050	0.674	0.075	0.698	385	380	375	0.2827	0.06046
0.630	0.006	0.027	0.679	0.050	0.704	0.075	0.728	350	345	341	0.3117	0.05484
0.650	0.007	0.028	0.702	0.053	0.729	0.080	0.751	328	324	320	0.3318	0.05151
0.670	0.007	0.028	0.722	0.053	0.749	0.080	0.774	309	305	301	0.3526	0.04848
0.710	0.007	0.028	0.762	0.053	0.789	0.080	0.814	276	272	269	0.3959	0.04318
0.750	0.008	0.030	0.805	0.056	0.834	0.085	0.861	247	244	241	0.4418	0.03869
0.800	0.008	0.030	0.855	0.056	0.884	0.085	0.911	217	215	212	0.5027	0.03401
0.850	0.009	0.032	0.909	0.060	0.939	0.090	0.968	193	190	188	0.5675	0.03012
0.900	0.009	0.032	0.959	0.060	0.989	0.090	1.018	172	170	168	0.6362	0.02687
0.950	0.010	0.034	1.012	0.063	1.044	0.095	1.074	154	153	151	0.7088	0.02412
1.000	0.010	0.034	1.062	0.063	1.094	0.095	1.124	139	138	137	0.7854	0.02176
1.060	0.011	0.034	1.124	0.065	1.157	0.098	1.188	124	123	122	0.8825	0.01937
1.120	0.011	0.034	1.184	0.065	1.217	0.098	1.248	111	110	109	0.9852	0.01735
1.180	0.012	0.035	1.246	0.067	1.279	0.100	1.311	100	99	99	1.094	0.01563
1.250	0.013	0.035	1.316	0.067	1.349	0.100	1.381	89	89	88	1.227	0.01393
1.320	0.013	0.036	1.388	0.069	1.422	0.103	1.455	80	80	79	1.368	0.01249
1.400	0.014	0.036	1.468	0.069	1.502	0.103	1.535	71	71	70	1.539	0.01110
1.500	0.015	0.038	1.570	0.071	1.606	0.107	1.640	62	62	61	1.767	0.009673
1.600	0.016	0.038	1.670	0.071	1.706	0.107	1.740	55	54	54	2.011	0.008502
1.700	0.017	0.039	1.772	0.073	1.809	0.110	1.844	49	48	48	2.270	0.007531
1.800	0.018	0.039	1.872	0.073	1.909	0.110	1.944	43	43	43	2.545	0.006718
1.900	0.019	0.040	1.974	0.075	2.012	0.113	2.048	39	39	38	2.835	0.006029
2.000	0.020	0.040	2.074	0.075	2.112	0.113	2.148	35	35	35	3.142	0.005441
2.120	0.021	0.041	2.196	0.077	2.235	0.116	2.272	31	31	31	3.530	0.004843
2.240	0.022	0.041	2.316	0.077	2.355	0.116	2.392	28	28	28	3.941	0.004338
2.360	0.024	0.042	2.438	0.079	2.478	0.119	2.516	25	25	25	4.374	0.003908
2.500	0.025	0.042	2.578	0.079	2.618	0.119	2.656	23	22	22	4.909	0.003482
2.650	0.027	0.043	2.730	0.081	2.772	0.123	2.811	20	20	20	5.515	0.003099
2.800	0.028	0.043	2.880	0.081	2.922	0.123	2.961	18.0	17.9	17.8	6.158	0.002776
3.000	0.030	0.045	3.083	0.084	3.126	0.127	3.166	15.7	15.6	15.5	7.069	0.002418
3.150	0.032	0.045	3.233	0.084	3.276	0.127	3.316	14.2	14.2	14.1	7.793	0.002193
3.350	0.034	0.046	3.435	0.086	3.479	0.130	3.521	12.6	12.5	12.5	8.814	0.001939
3.550	0.036	0.046	3.635	0.086	3.679	0.130	3.721	11.2	11.2	11.1	9.898	0.001727
3.750	0.038	0.047	3.838	0.089	3.883	0.134	3.926	10.0	10.0	10.0	11.04	0.001548
4.000	0.040	0.047	4.088	0.089	4.133	0.134	4.176	8.8	8.8	8.8	12.57	0.001360
4.250	0.043	0.049	4.341	0.092	4.387	0.138	4.431	7.8	7.8	7.8	14.19	0.001205
4.500	0.045	0.049	4.591	0.092	4.637	0.138	4.681	7.0	7.0	6.9	15.90	0.001075
4.750	0.048	0.050	4.843	0.094	4.891	0.142	4.936	6.3	6.2	6.2	17.72	0.0009646
5.000	0.050	0.050	5.093	0.094	5.141	0.142	5.186	5.7	5.6	5.6	19.63	0.0008706

Cu wire dia millimetres	Overall dia, including enamel, mm	Cu wire dia millimetres	Overall dia, including enamel, mm	Cu wire dia millimetres	Overall dia, including enamel, mm
4.000	4.160	0.950	1.041	0.280	0.334
3.750	3.905	0.900	0.990	0.265	0.312
3.550	3.702	0.850	0.937	0.250	0.301
3.350	3.498	0.800	0.885	0.236	0.285
3.150	3.294	0.750	0.832	0.224	0.272
3.000	3.142	0.710	0.790	0.212	0.258
2.800	2.938	0.670	0.749	0.200	0.245
2.650	2.784	0.630	0.706	0.190	0.234
2.500	2.631	0.600	0.675	0.180	0.222
2.360	2.488	0.560	0.632	0.170	0.211
2.240	2.366	0.530	0.601	0.160	0.199
2.120	2.243	0.500	0.569	0.150	0.188
2.000	2.120	0.475	0.543	0.140	0.176
1.900	2.018	0.450	0.516	0.132	0.167
1.800	1.916	0.425	0.489	0.125	0.159
1.700	1.813	0.400	0.462	0.112	0.143
1.600	1.711	0.375	0.436	0.100	0.129
1.500	1.608	0.355	0.414	0.090	0.117
1.400	1.506	0.335	0.393	0.080	0.105
1.320	1.423	0.315	0.371	0.071	0.095
1.250	1.351	0.300	0.355	0.063	0.085
1.180	1.279	Metric winding wire sizes 200C polyester-imide GRADE 2		0.060	0.081
1.120	1.217			0.056	0.076
1.060	1.155			0.050	0.068
1.000	1.093				

5.4 THE OUTPUT TRANSFORMER

- Separates the high DC voltage of the output tube circuit from the load (loudspeaker).
- Matches the impedance of the load (loudspeaker) with the impedance of the tube.
- Reflects the load impedance of the Secondary back to the Primary.
- Transfers power from the output tube to the load (loudspeaker).
- The Primary of the output transformer is part of the output tube circuit and DC current from the power supply flows through it.

Good performance output transformer:

1. Wide frequency band.
2. Low distortion.
3. Low losses.

Electric diagram of the simplified equivalent circuit of the output transformer**:**

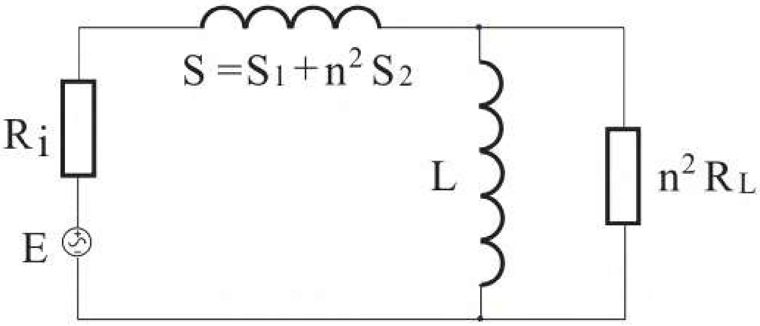


Fig. 5-04

R_i – Internal resistance of the source
(internal resistance of the tube)
R_a – Load Resistance of the Anode (Plate)
L – Inductance of the Primary
S – Leakage inductance
R_L – Load Resistance of the Secondary

Low frequencies:

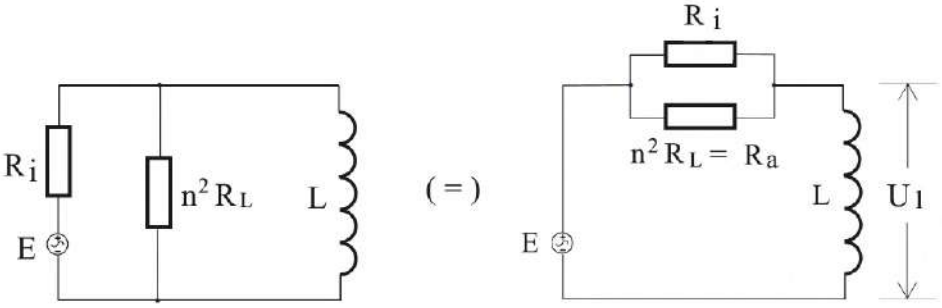


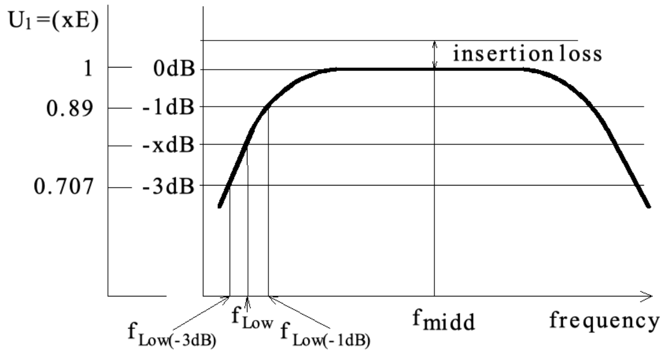
Fig. 5-05

Voltage attenuation at low frequencies

$$b_L = 20 \times \log \sqrt{1 + \left[\frac{R_i \times R_a}{(R_i + R_a) \times 2 \times \pi \times f \times L} \right]^2} \, [dB] \tag{501.20}$$

At the lower end of the frequency band (cut-off frequency $f_{Low(-3dB)}$ i.e. the frequency at which the voltage across the Primary of the transformer is 0.707 of the voltage value at middle frequencies), the inductance L is:

$$L = \frac{R_i \times R_a}{R_i + R_a} \times \frac{1}{2 \times \pi \times f_{Low(-3dB)}} \tag{501.21}$$



The Inductance **L** [H] for the attenuation **b** [dB] at the frequency **f** [Hz] is:

$$L = \frac{1}{2 \times \pi \times f} \times \frac{R_i \times R_a}{R_i + R_a} \times \frac{1}{\sqrt{10^{0.1 \times b} - 1}}$$

$$L = \frac{R_i \times R_a}{R_i + R_a} \times \frac{x}{2 \times \pi \times f_L} \qquad (501.22)$$

Inductance L for attenuation ***– 1 dB*** *(* $U_l = 0.89 \times E$ *) at frequency f (* ***x = 2****) is:*

$$L = \frac{R_i \times R_a}{R_i + R_a} \times \frac{2}{2 \times \pi \times f_L} = \frac{\boldsymbol{R_i \times R_a}}{\boldsymbol{R_i + R_a}} \times \frac{\mathbf{1}}{\boldsymbol{\pi \times f_L}}$$

Inductance L for attenuation ***– 0.5 dB*** *(* $U_l = 0.94 \times E$ *) at frequency f (* ***x = 2.86****) is:*

$$L \approx \frac{R_i \times R_a}{R_i + R_a} \times \frac{\mathbf{3}}{\boldsymbol{2 \times \pi \times f_L}}$$

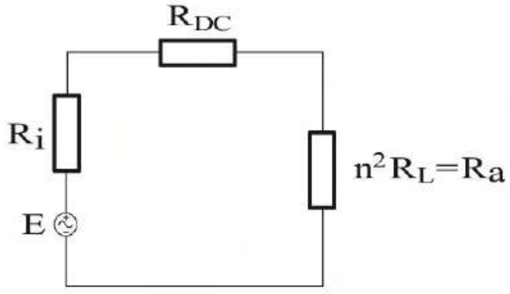
Inductance L for attenuation ***– 0.2 dB*** *(* $U_l = 0.977 \times E$ *) at frequency f (* ***x = 4.6****) is:*

$$L \approx \frac{R_i \times R_a}{R_i + R_a} \times \frac{\mathbf{4.6}}{\boldsymbol{2 \times \pi \times f_L}}$$

Inductance L for attenuation ***– 0.1 dB*** *(* $U_l = 0.988 \times E$ *) at frequency f (* ***x = 6.55****) is:*

$$L \approx \frac{R_i \times R_a}{R_i + R_a} \times \frac{\mathbf{1}}{\boldsymbol{f_L}}$$

Middle frequencies



$\mathbf{R_{DC} = R_{P\,DC} + R_{S\,DC}}$

$R_{P\,DC}$ – DC resistance of the Primary coil
(resistance of the Cu wire of the Primary coil)

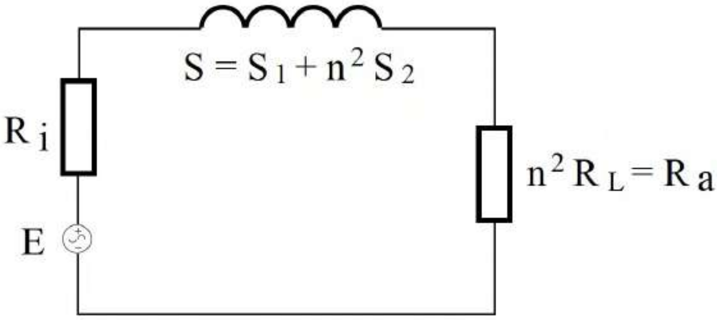
R_{Sdc} – DC resistance of the Secondary coil
(resistance of the Cu wire of the Secondary coil)

The voltage across the Primary is lower than the generator voltage due the voltage drop on the internal resistance of the tube and the voltage drop on the DC resistance of the transformer coil.

Voltage attenuation at middle frequencies

$$\boldsymbol{b_M} = 20 \times log\left(1 + \frac{R_{DC}}{R_i + R_a}\right) [dB] \qquad (501.23)$$

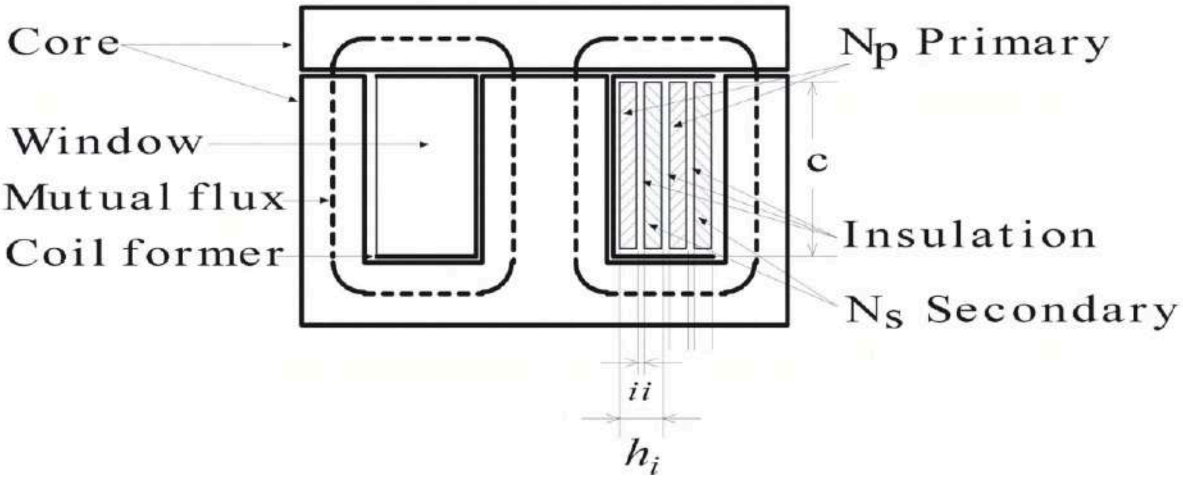
High frequencies



N_p – number of turns of the Primary
n_d – number of dielectrics, i.e., the number of physical contacts between the coil layers of the Primary and the Secondary
d_g [mm] – the dielectric gap, i.e., the distance between the surfaces of the cooper wire of the Primary and Secondary turns
h_0 [mm] – total radial dimension of the coils
c″ [mm]– the traverse width of the coil

S – Leakage inductance

Leakage inductance is an unwanted effect caused by an imperfect link between the coils.
The magnetic flux that does not link the Primary coil to the Secondary coil acts as an inductivity connected in series with the Primary coil.
The leakage inductance with unwanted capacitances between the transformer coils forms a high-pass filter which affects the frequency response of the amplifier at the higher end of the frequency band.
The leakage inductance depends on the geometry of the coils and the number turns of the coils.
To minimize these unwanted effects, the Primary coil is divided into a number of sections, and the Secondary coils (sections) are usually inserted between the sections of the Primary coil.



c – coil length – (mm)
h_i – coil height (mm)
ii – insulation thickness (mm)

Leakage inductance of split Primary coil and inserted Secondary coils:

$$S = \frac{0.417 \times N_P^2 \times \ell_t \times [(2 \times n_d \times d_g) + h_0]}{10^9 \times n_d^2 \times c}\ [H] \qquad (501.24)$$

N_p – number of turns of the Primary
n_d – number of dielectrics, i.e., the number of physical contacts between the coil layers of the Primary and the Secondary

d_g[mm] – the dielectric gap, i.e. the distance between the surfaces of the cooper wire of the Primary and Secondary turns
h_0[mm] – total radial dimension of the coil
c[mm] – the traverse width of the coil
ℓ_t[mm] – mean length per turn (MLT)

Voltage attenuation at high frequencies

$$b_H = 20 \times log\sqrt{1+\left[\frac{2\times\pi\times f\times S}{(R_i+R_a)}\right]^2}\ [dB] \qquad (501.24)$$

Higher end of the frequency band – cut-off frequency $f_{High(-3dB)}$:

$$f_{High(-3dB)} = \frac{R_i+R_a}{2\times\pi\times S} \qquad (501.25)$$

 # The high cut-off frequency is also determined by the stray capacitances (the capacitance between the Primary windings and the earthed windings of the Secondary and the self-capacitances between adjacent turns of the Primary coils.

For a typical output transformer design, this capacitance has the value: 350 pF ≤ C_{sh} ≤ 1300 pF.

Leakage inductance S transferred to the Primary can be calculated by measuring the impedance of the Primary under condition of short-circuited Secondary:

$$S = \frac{1}{2\times\pi\times f}\sqrt{Z^2-(R_1+n^2\times R_2)^2} \qquad (501.26)$$

Z – measured impedance of the Primary under the above conditions
R_1 – DC resistance of the Primary
n^2 x R_2 – impedance of the Primary

- A -

The characteristics of the output transformer at high frequencies mostly depend on the geometry of the coils and the number of turns of the coils (leakage inductance, stray capacitances) and the magnetic characteristics of the core laminate material and the thickness of the laminate.

- B -

The characteristics of the output transformer at low frequencies mostly depend on the inductance of the Primary and the mass of the ferromagnetic core, i.e. cross-section of the core.

- C -

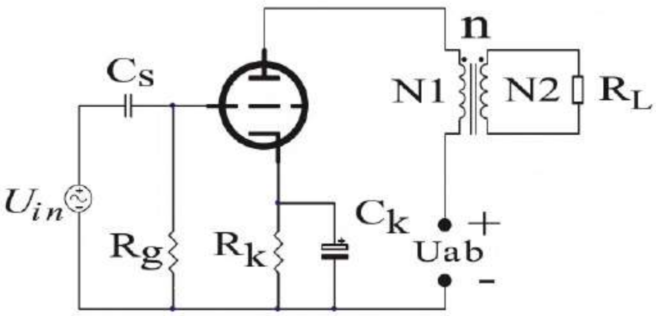
Distortions of the output transformer, especially at low frequencies, mostly depend on the inductance of the Primary and the cross-section of the core, i.e. the maximum flux density – it is necessary to use a linear part of the B – H curve.

- D -

The power of the output transformer is determined by the size of the core and the size of the wire.

5.5 CALCULATION OF SINGLE-ENDED OUTPUT TRANSFORMER

Single-Ended Output stage basic configuration:



The Primary of the output transformer is part of the DC power supply of the output tube, so the DC anode current (anode quiescent current) also flows through the Primary and produces a magnetic field in the ferromagnetic core.

The calculation of the output transformer is an iterative process, i.e. it starts with some input data and if the calculation doesn't give acceptable results, calculation starts again.

Input data:

R_i *– Tube internal resistance*
I_a *– Plate (Anode) DC current at the operating point of the tube (QUISCENT Point)*
U_a *– Plate (Anode) DC Voltage at the quiescent point*
R_{La} *– Plate (Anode) Load*
P_{out} *– Output Power*
f_{min} *– The lowest frequency (Hz) at which the transformer can operate at full power*
f_{-3dB}* –** *cut-off frequencies of the bandwidth (* ***$f_{-3dB\ low}$ *or* ***$f_{-3dB\ high}$*** *)*

1. The first step of calculation is defining some necessary data

Define ***R_{La}*** *using the* ***LOAD Line,***
(Empirical choice: ***$R_{La} \geq 5 \times R_i$****) or use a quick calculation:*

$$R_{La} = \frac{U_{a(quiscent\ point)}}{I_{a(quiscent\ point)}} - 2 \times R_i$$

$$P_{0_{max}} = \frac{R_{La} \times I^2{}_{a(quiscent\ point)}}{2}$$

$$P_{0_{max(ac)}} = \frac{1}{2} \times \frac{(U_{a(quiscent\ point)} - R_i \times I_{a(quiscent\ point)})^2}{(R_i + R_{La})^2} \times R_{La}$$

Define required or desired ***L_a*** *and* ***f_{min}****:*

$$L_a = \frac{R_{La}}{2 \times \pi \times f_{low}}$$

$$R_A = \frac{(R_i + R_{Pdc}) \times R_{La}}{R_i + R_{Pdc} + R_{La}}$$

$$\frac{A_{Low\ f}}{A_{Mid\ f}} = \frac{1}{\sqrt{1 + (\frac{R_i + R_{La}}{2 \times \pi \times L_a})^2}}$$

$$f_{-3dB\ low\ f} = \frac{R_A}{2 \times \pi \times L_a}$$

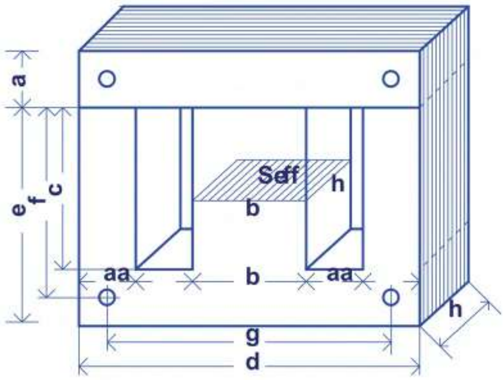
$$f_{-1dB\ low\ f} = \frac{R_A}{\pi \times L_a}$$

$$R_B = R_i + R_{Pdc} + R_{La}$$

$$f_{-3dB\ high\ f} = \frac{R_B}{2 \times \pi \times L_s}$$

***L_S* –** *Total Leakage Inductance of the Primary*

Calculation of the Iron Core Cross — Section (S_{Fe})



First method:

Equation of the electromotive force **(emf)** of the Primary coil:

$$\boldsymbol{emf_p} = 4.44 \times 10^{-4} \times f \times N_p \times B_m \times S_{eff}$$

Although the emf of the Primary and Secondary differ from the supply voltage U_p of the Primary and the induced voltage U_s of the Secondary (for voltage drops across the coils), $\boldsymbol{emf_p \approx U_p}$ *can be used:*

$$\boldsymbol{U_p} = 4.44 \times 10^{-4} \times f \times N_p \times B_m \times S_{eff}$$

If assume that the Primary coil occupies half of the core window (secondary coil occupies half of the core window), and the Cu wire filling coefficient is 0.25:

Cu cross-section of the Primary coil is:

$$N_p \times d_p = \frac{S_W}{2} \times 0.25 \times 100 = 12.5 \times S_W$$

Where: N_p – number of turns of the Primary coil; d_p – diameter of the Cu wire of the Primary coil; S_W – Cross-section of the transformer core window

Electric current of the Primary coil:

$$\boldsymbol{I_p} = d_p \times J, \qquad \boldsymbol{or} \quad \boldsymbol{d_p} = \frac{I_p}{J_p}$$

J – Electric current density

Substituting $\boldsymbol{d_p} = \frac{I_p}{J_p}$ into the above equation:

$$N_p \times d_p = N_p \times \frac{I_p}{J_p} = \frac{S_W}{2} \times 0.25 \times 100 = 12.5 \times S_W$$

Or

$$\boldsymbol{I_p} = \frac{12.5 \times S_W \times J}{N_p}$$

Electric power of the Primary: $\boldsymbol{P_p = U_p \times I_p}$:

$$\boldsymbol{P_p} = (4.44 \times 10^{-8} \times f \times N_p \times B \times S_{eff}) \times \frac{12.5 \times S_W \times J}{N_p}$$

$$\boldsymbol{P_p} = 55.5 \times 10^{-8} \times f \times B \times S_{eff} \times S_W \times J$$

From the equation above, the cross-section of the iron core is:

$$\boldsymbol{S_{Fe}}(cm^2) = \frac{1.8 \times 10^6 \times P_p(W)}{f(Hz) \times B(gauss) \times S_W(cm^2) \times J(\frac{A}{mm^2})} \qquad (501.27)$$

B (Gauss) – Magnetic Flux Density
f (Hz) – The lowest frequency (Hz) at which the transformer can operate at full power
J (A/mm^2) – Electric Current Density
P (W) – Transformer power
S_W (cm^2) – Cross-section of the transformer core window (For E/I laminate S_W = (a x c) [cm^2]
Calculation starts with some standard type of E/I transformer laminate, i.e. S_W .

If the proportion <u>$1 \leq h / b \leq 1.6$</u> is fulfilled, the choice of transformer laminate is good, but if the mentioned proportion is not fulfilled, the calculation process must be repeated with a new standard transformer laminate (S_W).

The advantage of the above method of calculating the cross-section (S_{Fe}) of the iron core is that there is a high probability that the coils of wire can be placed in the coil former.

The above equation can be simplified using numerical values (**good performance output transformer)**, **B = 5000 Gauss** (relatively constant inductance), **J = 1.5A/mm^2** , and f a little lower than 20Hz (**f ≈ (14 ÷ 18) Hz**):

$$\boldsymbol{S_{Fe}(cm^2)} \approx \frac{(14 \div 17) \times P(W)}{S_W(cm^2)} \qquad (501.28)$$

* * *

** Mathematical derivation of the above equation*

Starting with the equation for the energy density in the magnetic material: E = H x B, and multiplying it by the volume of the core ($V_{Fe}=S_{Fe}x\ell_{Fe}$) and assuming a sinusoidal form of B at frequency f, the equation can be written as follows:

$$P(VA) = 4.44 \times l_{Fe} \times S_{Fe} \times f \times B \times H \times 10^{-8}$$

Considering that $H = \frac{N \times I}{l_{Fe}}$ and $N \times I = k \times J \times S_w$, *(k – filling coefficient, 0.25 for Primary and Secondary windings), the transformer power equation can be written as:*

$$P(VA) = 0.555 \times J\left(\frac{A}{mm^2}\right) \times B(Gauss) \times f(Hz) \times S_{Fe}(cm^2) \times S_W(cm^2) \times 10^{-6}$$

Also, starting the derivation of the equation from the Faraday's Low of Induction ($emf = -\frac{\Delta\phi}{\Delta t} = -N \times S_{Fe} \times \frac{dB}{dT}$ *) and solving the equation for the case of sinusoidal flux:*

$$S_{Fe} = \frac{1.8 \times P(W) \times 10^6}{S_w(cm^2) \times J(\frac{A}{mm^2}) \times B(Gauss) \times f(Hz)}$$

* * *

Example

Designing a SE output transformer (output tube: WE 300B):
P_{out} = 8.5 W
R_{La} = 3500 Ω
$L_{a(min)}$ = 24.3 H (95% of full power or power attenuation of – 0.22 dB at 18 Hz)
WE 300B: R_i = 720 Ω, I_a = 81 mA, U_a = 350 V, $U_{G\text{-}K}$ = 71 V

First attempt:

*Calculation attempt with standard **EI 108** transformer laminate:*
a = 1.8 cm
b = 3.6 cm
c = 3 x a = 3 x 1.8 cm = 5.4 cm
S_W = a x c = 1.8 x 5.4 cm = 9.72 cm^2

$$\boldsymbol{S_{Fe}}(cm^2) = \frac{17 \times \boldsymbol{P}(W)}{\boldsymbol{S_W}(cm^2)} = \frac{17\ x\ 8.5(w)}{9.72(cm^2)} = \mathbf{14.87\ cm^2}$$

$$h(cm) = \frac{S_{Fe}(cm^2)}{b(cm)} = \frac{14.87}{3.6} = \mathbf{4.13\ cm}$$

Second attempt:
The calculation can be repeated with the next larger standard transformer laminate: EI 120
a = 2.0 cm
b = 4.0 cm
c = 3 x a = 3 x 2.0 cm = 6.0 cm
$S_W = a \times c = 2.0\ cm \times 6.0\ cm = 12.0\ cm^2$

$$\boldsymbol{S_{Fe}}(cm^2) = \frac{17 \times \boldsymbol{P}(W)}{\boldsymbol{S_W}(cm^2)} = \frac{17\ x\ 8.5(W)}{12(cm^2)} = \mathbf{12.04\ cm^2}$$

$$h(cm) = \frac{S_{Fe}(cm^2)}{b(cm)} = \frac{12.04}{4} = 3.01\ cm$$

Second method:

$$\boldsymbol{S_{Fe(min)}(cm^2)} = \sqrt{\frac{P_{0_{maxac}} \times G \times 10^6}{B \times J \times f}} \tag{501.29}$$

$\boldsymbol{P_{0maxac}}$ **–** *Output power [W]*
B – *Induction* ***(4000 ÷ 8000) Gs*** *for transformer laminate of 4% Si Ferro alloy*
G – *Coefficient of weight ratio of iron core and cooper winding* ***(1.7 to 2.5)***
J – *Current density* ***(1.5 to 2.5) A / mm²***
f – *The lowest frequency (Hz) at which the transformer can operate at full power*

The above equation can be simplified using B = 5000 Gauss, J = 1.5A/mm^2 and G = 2.5 (good performance output transformer):

$$\boldsymbol{S_{Fe}(cm^2)} = 18.26 \times \sqrt{\frac{P_{0_{maxac}}}{f_{min}}}$$

and for f_{min} a little lower than 20Hz (14 ÷ 18Hz):

$$\boldsymbol{S_{Fe}(cm^2) \approx (4.3 \div 4.9) \times \sqrt{P_{0_{max}}(W)}} \tag{501.30}$$

The disadvantage of the above method of calculating the cross-section (S) of the iron core is that it <u>does not take into account</u> whether the coils of wire can fit into the coil former.

****Note:***

> *For the calculation of lower power output transformers (up to the 10W) with high impedance of the Primary, the* ***maximum*** *value of the* ***constant*** *in front of the square root of the above equation should be used:*

$$S_{Fe}(cm^2) \approx 4.9 \times \sqrt{P_{0_{max}}(W)}$$

Applied to the example above:

$$S_{Fe}(cm^2) \approx 4.9 \times \sqrt{P_{0_{max}}(W)} = 4.9 \times \sqrt{8.5} = 14.285 cm^2$$

Useful information that can be used to choose a standard transformer laminate:

$$h(cm) = b(cm) = \sqrt{S_{Fe(cm^2)}} = \sqrt{14.285} = 3.78\ cm$$

<u>CHOICE OF IRON CORE (STANDARD)</u>

Choice of standard transformer laminate.

Choice of standard (or preferred) coil former.

Standard dimension of EI laminates tongue (b) and the height of the laminates stack (h)

$$S_{Fe(st)} = b \times h \quad (cm^2)$$

The optimal choice is: b = h; (b / h) = 1

It is recommended to fulfill the ratio $1 \leq h / b \leq 1.6$.

It is recommended to choose b and h so that the condition $S_{Fe(st)} > S_{Fe}$ is fulfilled.

Calculation of the effective cross-section of the iron core $S_{Fe_{eff}}$.

$$S_{Fe_{eff}} = (0.92 - 0.98) \times (b \times h) \; [cm^2] \tag{501.31}$$

Stacking coefficient of 4% Si non-oriented steel (NOSS) transformer laminate:

1. thickness **0.35** mm **k = 0.96**
2. thickness **0.5** mm **k = 0.97**

Using the first method of calculation:

First attempt:
EI 108 *transformer laminate (b = 3.6 cm and h = 4.13 cm);* $S_{Fe} = 14.87 cm^2$

$$\boldsymbol{h} = \frac{S_{Fe}}{0.96 \times b} = \frac{14.87}{0.96 \times 3.6} = 4.3(cm^2)$$

*The nearest standard coil former: (**3.6 x 4.5**) cm.*

$S_{Fe\ eff}$ = 0.96 x (3.6 x 4.5) = **15.55 cm²**

Checking the inequality **$1 \leq h / b \leq 1.6$: $h / b = 4.5 / 3.6 = 1.25$.** *Condition $1 \leq h / b \leq 1.6$ is* ***FULFILLED.***

Second attempt:
EI 120 *transformer laminate (b = 4 cm and h = 3.01 cm);* $S_{Fe} = 12.04 cm^2$

$$\boldsymbol{h} = \frac{S_{Fe}}{0.96 \times b} = \frac{12.04}{0.96 \times 4} = \mathbf{3.13\ (cm^2)}$$

*The nearest standard coil former: (**4 x 3**) or (4 x 4) cm.*

$S_{Fe\ eff}$ = 0.96 x (4 x 3.0) = **11.52 cm²**

Checking the inequality **$1 \leq h / b \leq 1.6$: $h / b = 3 / 4 = 0.75$.** *Condition $1 \leq h / b \leq 1.6$ is* ***NOT FULFILLED.***

and: $S_{Fe\ eff}$ = 0.96 x (4 x 4.0) = **15.36 cm²**

Checking the inequality **$1 \leq h / b \leq 1.6$: $h / b = 4/ 4 = 1$.** *Condition $1 \leq h / b \leq 1.6$ is* ***FULFILLED .***

Using the second method of calculation:

$S_{Fe} = 14.285\ cm^2$
h = **3.57 cm**.
The nearest standard transformer laminate is ***EI 120 (b = 4.0 cm).***

$$\boldsymbol{h} = \frac{S_{Fe}}{0.96 \times b} = \frac{14.285}{0.96 \times 4} = \boldsymbol{3.72\ (cm^2)}$$

The nearest (larger size) standard coil former (***b** = **4*** *cm and* ***h = 4*** *cm).*

The effective cross-section of the transformer core is:

$$\boldsymbol{S_{Fe_{eff}}} = 0.96 \times (b \times h) = 0.96 \times (4 \times 4) = \boldsymbol{15.36\ (cm^2)}$$

Checking the inequality **1 ≤ h / b ≤ 1.6:**

h / b = 4 / 4 = 1. *Condition 1 ≤ h / b ≤ 1.6 is* ***FULFILLED.***

Conclusion of the calculation of the **cross-section** of the iron core:

1. EI transformer laminate **EI 108: Coil Former (3.6 x 4.5) cm; $S_{Fe\ eff}$ = 15.55 cm^2**
2. EI transformer laminate **EI 120: Coil Former (4 x 3.0) cm; $S_{Fe\ eff}$ = 11.52 cm^2 - ELIMINATED**
3. EI transformer laminate **EI 120: Coil Former (4 x 4) cm; $S_{Fe\ eff}$ = 15.36 cm^2**

The number of turns of the Primary:

First method:

Using the equations:

$$\boldsymbol{U_p} = 4.44 \times 10^{-4} \times f \times N_p \times B_m \times S_{eff}$$

and

$$\boldsymbol{U_P} = \sqrt{R_{La} \times P}$$

P – Output power (W)

R_{La} – Anode load (Ω) (impedance of the Primary of the output transformer)

$$\boldsymbol{N_p} = \frac{10^8 \times \sqrt{P \times R_{La}}}{4.44 \times f_{min} \times B_{max} \times S_{Fe_{eff}}} \qquad (501.32)$$

A simplified equation for calculating the number of turns of the Primary of ***higher power*** *SE output transformer and* ***lower Cu losses*** *of the Primary windings (4% Si GOSS transformer laminate B_{AC}+ B_{DC} = 8000 Gauss + 8000 Gauss = 16000 Gauss = 1.6 T):*

$$\boldsymbol{N_P} \approx \frac{(200 \div 230) \times \sqrt{P \times \mathrm{R_{La}}}}{S_{Fe_{eff}}} \qquad (501.33)$$

A simplified equation for calculating the number of turns of the Primary of ***lower power*** *SE output transformer and* ***higher Cu losses*** *of the Primary windings (4% Si NOSS transformer laminate B_{AC} + B_{DC} = 6000 Gauss+6000 Gauss = 12000 Gauss = 1.2 T):*

$$\boldsymbol{N_P} \approx \frac{(230 \div 300) \times \sqrt{P \times \mathrm{R_{La}}}}{S_{Fe_{eff}}} \qquad (501.34)$$

Example (cont.):
Transformer laminate: 4% Si NOSS
B_{max} = 5000 Gauss
f = 18 Hz

The minimum number of turns of the Primary:

1. *EI transformer laminate EI 108: Coil Former (3.6 x 4.5) cm;* $S_{Fe\ eff}$ *= 15.55 cm*^{2}

$$\boldsymbol{N_{P(\min)}} = \frac{10^8 \times \sqrt{P \times R_{La}}}{4.44 \times f_{min} \times B_{max} \times S_{Fe_{eff}}} = \frac{10^8 \times \sqrt{8.5W \times 3500\Omega}}{4.44 \times 18Hz \times 5000 \times 15.55cm^2} = \boldsymbol{2776\ turns}$$

2. *EI transformer laminate EI 120: Coil Former (4 x 4) cm;* $S_{Fe\ eff}$ *= 15.36 cm*^{2}

$$\boldsymbol{N_{P(\min)}} = \frac{10^8 \times \sqrt{P \times R_{La}}}{4.44 \times f_{min} \times B_{max} \times S_{Fe_{eff}}} = \frac{10^8 \times \sqrt{8.5W \times 3500\Omega}}{4.44 \times 18Hz \times 5000 \times 15.36cm^2} = \boldsymbol{2810\ turns}$$

Second method:

Assuming that the Primary and Secondary windings each occupy half of the core window and that the Cu wire filling coefficient is: 0.25 (max. 0.35):

$$\boldsymbol{N_P} = 12.5 \times S_w \times \frac{J}{I} \tag{501.35}$$

J – Electric current density (A / mm^2) (***1.5 A / mm^2 ≤ J ≤ 2.5 A / mm^2***)
I - DC current (A) flowing through the Primary winding

For J = 2.5 A / mm^2 : $\boldsymbol{N_P \approx 30 \times \frac{S_w}{I}}$

For J = 2.0 A / mm^2 : $\boldsymbol{N_P \approx 25 \times \frac{S_w}{I}}$

For J = 1.5 A / mm^2 : $\boldsymbol{N_P \approx 18.75 \times \frac{S_w}{I}}$

I (A) – DC current (A) flowing through the Primary winding
S_W (cm^2) – Cross-section of the transformer core window *(For E/I laminate:* S_W *= (a x c) [cm*^{2}*])*

A lower DC current density (J) means a lower number of turns of the Primary winding, but also a larger diameter of the wire of the Primary winding and lower Cu losses of the Primary.

A simplified equation for calculating the number of turns of the Primary of Hi End SE output transformer with low Cu losses of the Primary:

$$\boldsymbol{N_P} \approx (20 \div 25) \times \frac{S_w}{I} \tag{501.36}$$

Example (cont.):

1. *EI transformer laminate EI 108: Coil Former (3.6 x 4.5) cm;* $S_{Fe\ eff}$ *= 15.55 cm*^{2}*;*
   Cross- section of transformer core window: S_W *= (a x c) = (1.8 x 5.4) = 9.72 cm*^{2}

   J = ***1.5 A/mm^2***

$$\boldsymbol{N_P} = 12.5 \times S_w \times \frac{J}{I} = 12.5 \times 9.72cm^2 \times \frac{1.5\frac{A}{mm^2}}{0.081A} = \boldsymbol{2250\ turns}$$

2. *EI transformer laminate EI 108: Coil Former (3.6 x 4.5) cm;* $S_{Fe\ eff}$ *= 15.55cm*^{2}*;*
   Cross-section of transformer core window: S_W *= (a x c) = (1.8 x 5.4) = 9.72 cm*^{2}

   J = ***2 A/mm^2***

$$\boldsymbol{N_P} = 12.5 \times S_w \times \frac{J}{I} = 12.5 \times 9.72cm^2 \times \frac{2\frac{A}{mm^2}}{0.081A} = \boldsymbol{3000\ turns}$$

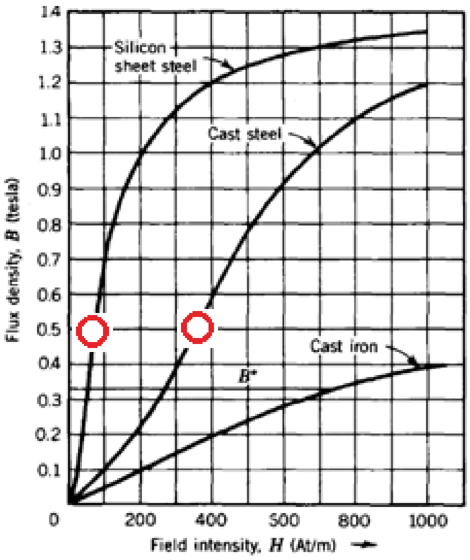
3. *EI transformer laminate EI 120: Coil Former (4 x 4) cm; $S_{Fe\ eff}$ = 15.36 cm^2;*
   Cross-section of transformer core window: S_W = (a x c) = (2 x 6) = 12 cm^2

   J = **1.5A /mm²**

$$N_P = 12.5 \times S_w \times \frac{J}{I} = 12.5 \times 12cm^2 \times \frac{1.5\frac{A}{mm^2}}{0.081A} = \boldsymbol{2777\ turns}$$

Third Method:

If the μ_e is ***defined*** *(the characteristics of the transformer laminate material are well known), the* ***permissible number*** *of* ***turns of the Primary*** *at* ***DC current*** $\boldsymbol{I_{DC}}$ *is:*



Flux (DC current) SE Quiescent Point

$$N_P = \frac{B_{DC} \times l_{Fe}}{1.257 \times \mu_e \times I_{DC}} \tag{501.37}$$

In a class A amplifiers, the peak AC current of the signal that flows through the anode load, i.e. through the output transformer, does not exceed the value of the DC current that constantly flows through the output transformer (quiescent current). Considering that the flux density is proportional to the current flowing through the windings of the transformer, a value equal to half of the maximum permissible flux density (the middle of the linear part of the B-H curve) is used for the DC flux density (B_{DC}), which avoids the operation of the transformer in the non-linear part (curved) part of the B-H curve close to saturation or even saturation of the transformer and provides low harmonic distortion.

In the process of designing and calculating the SE output transformer, as a precautionary measure, it is recommended to use B_{DC} = (4000 ÷ 5000) Gauss.

The above equation can be used to calculate μ_e and air gap dimension in the case where the iron core and the number of Primary turns are already defined.

In the process of designing and calculating the SE output transformer, it is recommended to use μ_e (effective permeability of 4% Si transformer laminate) in the range of:

NOSS: $200 \leq \mu_e \leq 450$ (Best: $250 \leq \mu_e \leq 350$)
GOSS: $200 \leq \mu_e \leq 750$ (Best: $250 \leq \mu_e \leq 450$)

Example (cont.):

1. *EI transformer laminate EI 108: Coil Former (3.6 x 4.5) cm; $S_{Fe\ eff}$ = 15.55 cm^2;*
   Cross-section of transformer core window: S_W = (a x c) = (1.8 x 5.4) = 9.72 cm^2
   Magnetic circuit length: MPL = $l_{Fe}(cm) = 2 \times c(cm) + 2 \times a(cm) + \pi \times a(cm)$

$$l_{Fe}(cm) = 2 \times 5.4 + 2 \times 1.8 + 3.14 \times 1.8 = 20.05\,cm$$

$\boldsymbol{\mu_e = 350}$
$\boldsymbol{B_{DC} = 5000}$ **Gauss**

$$N_P = \frac{B_{DC} \times l_{Fe}}{1.257 \times \mu_e \times I_{dc}} = \frac{5000 \times 20.05}{1.257 \times 350 \times 0.081} = 2813$$

2. *EI transformer laminate EI 120: Coil Former (4 x 4) cm; $S_{Fe\ eff}$ = 15.36 cm^2;*
   Cross-section of transformer core window: S_W = (a x c) = (2 x 6) = 12 cm^2
   Magnetic circuit length: MPL = $l_{Fe}(cm) = 2\times c(cm) + 2\times a(cm) + \pi\times a(cm)$

$$l_{Fe}(cm) = 2\times 6 + 2\times 2 + 3.14\times 2 = 22.28\,cm$$

μ_e = 350
B_{DC} = 5000 Gauss

$$N_P = \frac{B_{DC}\times l_{Fe}}{1.257\times \mu_e \times I_{dc}} = \frac{5000\times 22.28}{1.257\times 350\times 0.081} = 3126$$

<u>The number of turns of the Secondary:</u>

$$N_S = N_P \times \sqrt{\frac{R_S}{R_{La}}} \qquad (501.38)$$

N_P – The number of turns of the Primary
N_S – The number of turns of the Secondary
R_{La} – Primary load (Ω)
R_S – Secondary load (Ω)

Example (cont.):

First method:

1. *EI transformer laminate EI 108: Coil Former (3.6 x 4.5) cm; $S_{Fe\ eff}$ = 15.55 cm^2*
   N_P = 2776 Turns

$$N_S = N_P \times \sqrt{\frac{R_S}{R_{La}}} = 2776\times\sqrt{\frac{4\,\Omega}{3500\,\Omega}} = 93.84\,Turns$$

2. *EI transformer laminate EI 120: Coil Former (4 x 4) cm; $S_{Fe\ eff}$ = 15.36 cm^2*
   N_P = 2810 Turns

$$N_S = N_P \times \sqrt{\frac{R_S}{R_{La}}} = 3123\times\sqrt{\frac{4\,\Omega}{3500\,\Omega}} = 94.9\,Turns$$

Second method:

1. *EI transformer laminate EI 108: Coil Former (3.6 x 4.55) cm; $S_{Fe\ eff}$ = 15.55 cm^2*
   Cross-section of transformer core window S_W = (a x c) = (1.8 x 5.4) = 9.72 cm^2; J = 1.5 A/mm^2;
   N_P = 2250 Turns

$$N_S = N_P \times \sqrt{\frac{R_S}{R_{La}}} = 2250\times\sqrt{\frac{4\,\Omega}{3500\,\Omega}} = 76\,Turns$$

2. *EI transformer laminate EI 120: Coil Former (4 x 4) cm; $S_{Fe\ eff}$ = 15.36 cm^2*
   Cross-section of transformer core window S_W = (a x c) = (2 x 6) = 12 cm^2; J = 1.5 A/mm^2;
   N_P = 2777 Turns

$$N_S = N_P \times \sqrt{\frac{R_S}{R_{La}}} = 2777\times\sqrt{\frac{4\,\Omega}{3500\,\Omega}} = 96.8\,Turns$$

<u>Wire diameter of the Primary and Secondary windings:</u>

Wire diameter of the Primary winding:

$$d_{Pmin}(mm) = 1.13\times\sqrt{\frac{I_a(A)}{J(\frac{A}{mm^2})}} \qquad (501.39)$$

$I_{P(DC)}$ –*Plate (Anode) DC Current*

$$For\ J = 1.5\ \frac{A}{mm^2}:\quad \mathbf{d_{Pmin}(mm) \approx 0.9 \times \sqrt{I_{P(DC)}(A)}}$$

$$For\ J = 2\ \frac{A}{mm^2}:\quad \mathbf{d_{Pmin}(mm) \approx 0.8 \times \sqrt{I_{P(DC)}(A)}}$$

$$For\ J = 2.5\ \frac{A}{mm^2}:\quad \mathbf{d_{Pmin}(mm) \approx 0.7 \times \sqrt{I_{P(DC)}(A)}}$$

Choosing a standard wire diameter*:*

$d_{P(standard)} \geq d_{Pmin}$

Example (cont.):

$$For\ J = 1.5\ \frac{A}{mm^2}:\quad \mathbf{d_{Pmin}(mm)} \approx 0.9 \times \sqrt{I_{P(DC)}(A)} = 0.9 \times \sqrt{0.081(A)} = \mathbf{0.256\ mm}$$

$d_{P(standard)}$ = 0.265 mm ≥ d_{Pmin} = 0.256 mm; Isolated $d_{P(standard)}$ = 0.265 mm = $d_{Wp(standard)}$ = 0.312 mm

$$For\ J = 2\ \frac{A}{mm^2}:\quad \mathbf{d_{Pmin}(mm)} \approx 0.8 \times \sqrt{I_{P(DC)}(A)} = 0.8 \times \sqrt{0.081(A)} = \mathbf{0.227\ mm}$$

$d_{P(standard)}$ = 0.236 mm ≥ d_{Pmin} = 0.227 mm; Isolated $d_{P(standard)}$ = 0.236 mm = $d_{Wp(standard)}$ = 0.285 mm

$$For\ J = 2.5\ \frac{A}{mm^2}:\quad \mathbf{d_{Pmin}(mm)} \approx 0.7 \times \sqrt{I_{P(DC)}(A)} = 0.7 \times \sqrt{0.081(A)} = \mathbf{0.199\ mm}$$

$d_{P(standard)}$ = 0.200 mm ≥ d_{Pmin} = 0.199 mm; Isolated $d_{P(standard)}$ = 0.200 mm = $d_{Wp(standard)}$ = 0.245 mm;

Wire diameter of the Secondary winding:

I_S = Secondary Current

$$\mathbf{I_S} = \frac{N_P}{N_S} \times \sqrt{\frac{P}{R_{La}}} \tag{501.40}$$

$$\mathbf{1.5}\ \frac{A}{mm^2} \leq J \leq \mathbf{2.5}\ \frac{A}{mm^2}$$

$$\mathbf{d_{Smin}(mm)} = 1.13 \times \sqrt{\frac{I_S(A)}{J(\frac{A}{mm^2})}} \tag{501.41}$$

Example (cont.):

$$\mathbf{I_S} = \frac{N_P}{N_S} \times \sqrt{\frac{P}{R_{La}}} = \sqrt{\frac{R_{La}}{R_S}} \times \sqrt{\frac{P}{R_{La}}} = \sqrt{\frac{3500}{4}} \times \sqrt{\frac{8.5}{3500}} = \mathbf{1.457\ A}$$

$$For\ J = 1.5\ \frac{A}{mm^2}:\quad \mathbf{d_{Smin}(mm)} \approx 0.9 \times \sqrt{I_S(A)} = 0.9 \times \sqrt{1.457(A)} = \mathbf{1.08}\ \mathrm{mm}$$

$d_{S(standard)}$ = 1.12 mm ≥ d_{Smin} = 1.08 mm

$$For\ J = 2\ \frac{A}{mm^2}:\quad \mathbf{d_{Smin}(mm)} \approx 0.8 \times \sqrt{I_S(A)} = 0.8 \times \sqrt{1.457(A)} = \mathbf{0.965}\ \mathrm{mm}$$

$d_{S(standard)}$ = 1.00 mm ≥ d_{Smin} = 0.965 mm

$$For\ J = 2.5\ \frac{A}{mm^2}:\quad \boldsymbol{d_{Smin}(mm)} \approx 0.7 \times \sqrt{I_S(A)}\ = 0.7 \times \sqrt{1.457(A)}\ = \mathbf{0.845}\ \mathrm{mm}$$

$d_{S(standard)}$ = 0.85 mm ≥ d_{Smin} = 0.845 mm

* * *

The possibility of placing all transformer windings inside the window of the transformer core:

*The possibility of placing all the windings inside the window of the transformer core (taking into account insulation of the wire, the design of coil former, the insulating foil between the winding layers) can be checked by calculating the **filling coefficient**:*

$$filling\ coefficient\ (\boldsymbol{p}) = \frac{C_u\ cross-section\ of\ all\ windings}{window\ cross-section\ (Sw)\ of\ the\ core}$$

Cu cross-section of all windings is the total cross-section of the Cu wires of all transformer windings.

- Transformer windings made of Cu insulated wire and without insulating foils between the windings layers: *the filling coefficient must be less than or equal to **0.35**.*
- *Transformer windings made of Cu insulated wire and **with** insulating foils between the windings layers: the filling coefficient must be less than or equal to **0.3**.*

*When designing and calculating the output transformer, it is recommended to use a filling cofficient p = **0.3**, i.e. the cross-section of Cu wires of all windings (primary and secondary) divided by the cross-section of the window of the transformer core must be less than or equal to 0.3.*
The result of the calculation of the filling coefficient can also be used very usefully to check the results of the design calculation, i.e. whether is it even possible to physically realize (make) a transformer using the results of the calculation.

The maximum diameter of the Cu wire of the Primary and Secondary windings can be calculated using the filling coefficient:
Maximum diameter of the Cu wire of the Primary winding
Assuming that the Primary windings of the output transformer occupy 1 / 2 of the cross-section of the transformer core:

$$\frac{\frac{d_P^2}{4} \times \pi \times N_P}{\frac{1}{2} \times S_W(mm^2)} = 0.3$$

*the maximum diameter of the Cu wire of the Primary winding **d_P** is:*

$$\boldsymbol{d_P(mm)} = \sqrt{\frac{0.3 \times 4 \times \frac{1}{2} \times S_W(mm^2)}{\pi \times N_P}} = \sqrt{\frac{\mathbf{0.19} \times \boldsymbol{S_W(mm^2)}}{\boldsymbol{N_P}}}$$

**** An equation for calculating the maximum diameter of the <u>insulated</u> Cu wire (<u>grade 2</u>) of the Primary winding has been published in some technical publications, which is very suitable for practical application:***

$$\boldsymbol{d_{Wp}(mm)} = \sqrt{\frac{0.28 \times S_W(mm^2)}{N_P}} \qquad (501.42)$$

- **Choose the wire according to the standards (nominal conductor diameter - $d_{P(standard)}$) and overall diameter including the isolation - (d_{Wp}))**

- **It is necessary to check whether the condition $d_{P(standard)} > d_{Pmin}$ is fulfilled, as well as whether the condition that the diameter of the selected insulated wire is smaller than the diameter of the insulated wire calculated using the above equation is fulfilled.**

If the minimum wire diameter of the Primary winding is defined or known, the above equation can be used to calculate the maximum number of turns of the Primary winding that can be placed in the selected EI transformer core window, or, if the number of turns of the Primary winding is defined or known, the above equation can be used to calculate the maximum wire diameter of the Primary winding (need to check: the maximum wire diameter must be equal to or bigger than the calculated wire diameter for the selected current density J).

$$d_{Wp}(mm) = \sqrt{\frac{0.28 \times S_W(mm^2)}{N_P}} \quad \rightarrow \quad N_{P(max)} = \frac{0.28 \times S_W}{d_{Wp}^2}$$

◆

Example (cont.):

EI 108 transformer laminate (S_W = 9.72 cm^2 = 972 mm^2) and J = 1.5 A/mm^2, d_{Wp} = 0.312 mm:

$$N_{P(max)} = \frac{0.28 \times S_W}{d_{Wp}^2} = \frac{0.28 \times 972}{0.312^2} = \mathbf{2795\ Turns}$$

EI 120 laminate (S_W = 12.0 cm^2 = 1200 mm^2) and J = 1.5 A/mm^2, d_{Wp} = 0.312 mm:

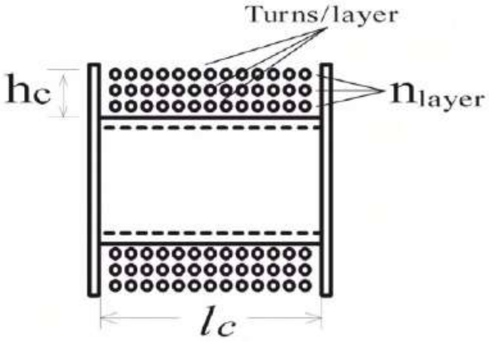
$$N_{P(max)} = \frac{0.28 \times S_W}{d_{Wp}^2} = \frac{0.28 \times 1200}{0.312^2} = \mathbf{3451\ Turns}$$

Calculating the number of turns of the Primary:

EI 108 transformer laminate, Coil former (3.6 x 4.5) cm, Effective cross-section of the transformer core $S_{Fe\ eff}$ =15.55cm^2
First method: ***N_P = 2776 Turns****, Second method:* ***N_P = 2250 Turns*** *and Third method:* ***N_P = 2813 Turns.***
The maximum number of turns of the Primary can be ***N_P = 2795 Turns***

EI 120 transformer laminate, Coil former (4 x 4) cm, Effective cross-section of the transformer core $S_{Fe\ eff}$ = 15.36 cm^2
First method: ***N_P = 2810 Turns****, Second method:* ***N_P = 2777 Turns*** *and Third method:* ***N_P = 3126 Turns.***
The maximum number of turns of the Primary can be ***N_P = 3451 Turns***

Calculation of turns per layer, number of layers, wire diameter, total number of winding turns



Effective lenght of one winding layer: $\ell_c = c'' = c - 4\ mm$

Number of turns per layer of the Primary:

$$Turn/Layer = \frac{0.97 \times c''}{d_{Wp(standard)}} = \frac{0.97 \times (c - 4mm)}{d_{Wp(standard)}\ (mm)} \qquad (501.43)$$

Choose the first lower INTEGER
(**sometimes it is possible to choose the first higher INTEGER)*

Number of layers of the Primary winding:

$$n_{layer} = \frac{N_P}{\left(\frac{Turns}{Layer}\right)} ;$$

$$\left({}^{Turn}/_{Layer} \right),\ n_{layer}\text{: } \underline{\textbf{INTEGER}}$$

Example (cont.):

EI 108 transformer laminate, Coil former (3.6 x 4.5) cm, Effective cross-section of the transformer core $S_{Fe\ eff} = 15.55 cm^2$
*The maximum number of turns of the Primary can be **$\underline{N_P = 2795}$ Turns***

$$\boldsymbol{{}^{Turn}/_{Layer}} = \frac{0.97 \times c''}{d_{Wp(standard)}} = \frac{0.97 \times (c - 4\,mm)}{d_{Wp(standard)}\ (mm)} = \frac{0.97 \times (54 - 4)}{0.312\ (mm)} = \mathbf{155.4}$$

*First **lower INTEGER**:* ${}^{Turn}/_{Layer} = 155$

$$n_{layer} = \frac{N_P}{\left(\frac{Turns}{Layer}\right)} = \frac{2795}{155} = \mathbf{18}$$

*n_{Layer} **INTEGER = 18***

Recalculated (new) number of turns of the Primary:

$$N_P = \left(\frac{Turns}{Layer}\right) \times n_{layer} = 155 \times 18 = \mathbf{2790\ Turns}$$

Recalculated (new) number of turns of the Secondary:

$$N_S = N_P \times \sqrt{\frac{R_S}{R_{La}}} = 2682 \times \sqrt{\frac{4\Omega}{3500\Omega}} = 94.3\ Turns;\ \textbf{INTEGER: 94 Turns}$$

EI 120 transformer laminate, Coil former (4 x 4) cm, Effective cross-section of the transformer core $S_{Fe\ eff} = 15.36 cm^2$
*The maximum number of turns of the Primary can be **$\underline{N_P = 3451}$ Turns***

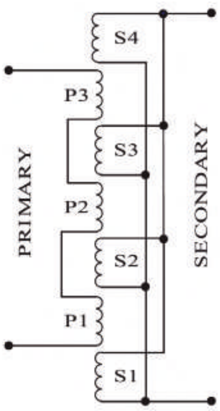
Primary: ${}^{Turn}/_{Layer}$ **= 172 ;** *n_{Layer} INTEGER* **= 20 ;** *N_P = **3440** Turns;* ***Secondary:*** *N_S = **116** Turns*

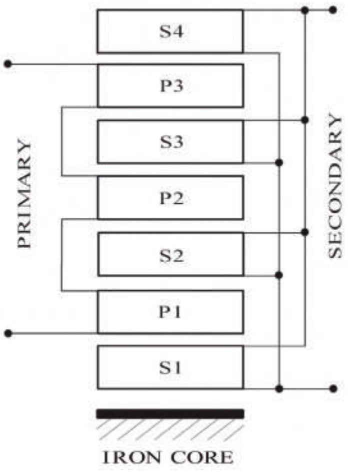
WINDING GEOMETRY – WINDING ARRANGEMENT

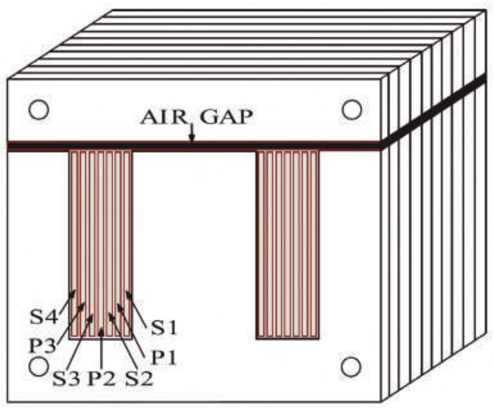
Design requirements:

- *Minimum Cu losses in the primary winding: **maximum wire diameter** i.e. minimum DC resistance of the winding.*
- *Maximum primary inductance i.e. minimum low cut-off frequency: **maximum number of turns of the primary (N_P)** taking into account the diameter of the wire from the aspect of placing the windings in the window of the transformer core.*
- *Minimum Leakage inductance i.e. maximu high cut-off frequency, optimum number of turns per layer: **winding geometry***

The leakage inductance (L_S) is highly dependent on the winding geometry and the number of winding turns.
The most commonly used method in practice to minimize leakage inductance as an unwanted effect is to divide the primary winding into sections and place the secondary windings between the primary sections. In practice, it also depends on the skill and experience of the designer. When deciding on the winding arrangement, the designer is usually guided by the number of layers of the primary winding and the power of the transformer.







Examples of **winding arrangement:**

Output transformers **of lower powers**

2S + 1P (two secondary sections and one primary section): (S – P – S)
2S + 3P (two secondary sections and three primary sections): (P/4 – S – P/2 – S – P/4)
3S + 4P (three secondary sections and four primary section): (P/6 – S – P/3 – S – P/3 – S – P/6)

Output transformer of **medium and higher power**

4S + 3P: (S – P/4 – S – P/2 – S – P/4 – S)
3S + 4P: [(3/20)P – S – (7/20)P – S – (7/20)P – S – (3/20)P]
4S + 5P: (P /8 – S – P/4 –S – P/4 – S – P/4 – S – P/8)
5S + 4P: [S - (4/18)P – S – (5/18)P – S – (5/18)P – S – (4/18)P – S]
…

Example (cont.):
EI 108 transformer laminate, Coil former (3.6 x 4.5) cm, $S_{Fe\ eff} = 15.55\ cm^2$
$N_P = 2790$ *Turns: 18 layers of 155 turns each. Insulated wire:* $d_{P(standard)} = 0.265\ mm$, $d_{Wp(standard)} = 0.312\ mm$
$N_S = 94$ *Turns*

Winding arrangement:
One of the simplest winding arrangements: the primary winding divided into three sections and two secondary sections placed between the primary sections (2S + 3P).

Primary section (4/18)P: 4 layers of wire winding – Secondary section* – Primary section (10/18)P: 10 layers of wire winding – Secondary section* – Primary section (4/18)P: 4 layers of wire winding

Secondary sections*:
$I_S = 1.457\ A$: *for* $J = 2\ A/mm^2$ *and* $1.5mm^2$ *it is necessary to use wire Ø 1.00 mm to Ø 1.12 mm respectively.*
EI 108 transformer laminate: the maximum lenght of one winding layer is 0.97 x (c – 4) mm = 0.97 x (54 – 4) = 48.5mm.
94 turns of wire Ø 1.00mm or Ø 1.12mm of the Secondary section cannot be wound in one layer (97 x 1mm > l_c *= 48.5mm). The problem can be solved, for example, by winding the Secondary section in two layers of (94 / 2) = 47 turns each.*
The wire diameter of a 47 turn winding that can be wounded in one layer on a EI 108 coil former is:

$d_{WP} = \frac{0.97\times c''}{N} = \frac{0.97\times(54mm-4mm)}{47} = 1.032\ mm$ *(max. diameter of insulated wire). Secondary section: two layers of*

47 turns each standard Cu wire Ø 0.9mm, insulated wire diameter $d_{WP} = 0.99mm$ *(0.99mm < max. wire diameter of a winding of 47 turns which can be wounded in one layer on a EI 108 coil former).*

If the output transformer option of $R_L = 8\ \Omega$ *is also required, an additional winding must be wound.*

Total number of turns of the 8Ω Secondary winding:

$$\boldsymbol{N_{S(8\Omega)}} = N_P \times \sqrt{\frac{R_S}{R_{La}}} = 2790 \times \sqrt{\frac{8\Omega}{3500\Omega}} = \mathbf{133\ \textit{Turns}}$$

The number of turns of the additional winding connected in series with the 4Ω winding to make an 8Ω Secondary is:
$N_{S(8\Omega)} - N_{S(4\Omega)} = 133 - 94 =$ ***39 Turns***
The wire diameter of a 39 turn winding that can be wounded in one layer on a EI 108 coil former is:
0.97 x (c - 4) mm /39 = 0.9 7x (54 - 4) /39 =1.24mm. Standard wire: Ø1.12mm (insulated wire diameter: 1.217mm).

The complete winding arrangement is:

- Primary section P1: 4 layers of 155 turns each (total: 620 turns), wire: Cul Ø 0.265mm, insulation foil between winding layers: 0.05mm
- Insulation foil: (0.3 ÷ 0.4) mm
- Secondary section $S1_{4\Omega}$: 2 layers of 47 turns each (total: 94 turns), wire: Cul Ø 0.9mm, insulation foil between winding layers: 0.05mm
- Insulation foil: 0.05mm
- Additional secondary section $S1_{add}$: 1 layer of 39 turns, wire: Cul Ø 1.12mm
- Insulation foil: (0.3 ÷ 0.4) mm
- Primary section P2: 10 layers of 155 turns each (total: 1550 turns), wire: Cul Ø 0.265mm, insulation foil between winding layers: 0.05mm
- Insulation foil: (0.3 ÷ 0.4) mm
- Secondary section $S2_{4\Omega}$: 2 layers of 47 turns each (total: 94 turns), wire: Cul Ø 0.9mm, insulation foil between winding layers: 0.05mm
- Insulation foil: 0.05mm
- Additional secondary section $S2_{add}$: 1 layer of 39 turns, wire: Cul Ø 1.12mm
- Insulation foil: (0.3 ÷ 0.4) mm
- Primary section P3: 4 layers of 155 turns each (total: 620 turns), wire: Cul Ø 0.265mm , insulation foil between winding layers: 0.05mm

Recommended thicknesses of insulation foils:
Insulation foil between the layers of the primary winding: $\boldsymbol{\Delta_i = 0.05mm}$
Insulation foil between the layers of the secondary winding: $\boldsymbol{\Delta_i = 0.05mm}$
Insulation foil between the layers of the primary and secondary windings: $\boldsymbol{\Delta_{ii} = (0.3 \div 0.4)\ mm}$

Calculation of the ***total radial dimension*** *of the windings including the insulation between the layers of the primary windings, between the layers of the secondary windings and between the layers of the primary and secondary windings:*

$$\boldsymbol{height_{total}} = n_{Players} \times (d_{P\,with\,isolation} + \Delta_i) + n_{Slayers} \times (d_{S\,with\,isolation} + \Delta_i) + n_{P-S\,layers} \times \Delta_{ii}$$

$$\boldsymbol{height_{total}} = \boldsymbol{h_0} = 18 \times (0.312 + 0.05) + 4 \times (0.99 + 0.05) + 2 \times (1.217 + 0.05) + 4 \times 0.4 = \mathbf{13.91}\boldsymbol{mm}$$

 # Condition: ***must be fulfilled:*** *13.91 ≤ 18 – 2.5 = 15.5 mm*

Calculation of ***mean length per turn (ℓ_t)****:*

$$\ell_t = 2 \times h + \pi \times h_0 + 2 \times b + 4 \times s = 2 \times 45 + 3.14 \times 13.91 + 2 \times 36 + 4 \times 2 = 213.6mm = 0.213m$$

Wire length of the primary winding (all sections): $\ell_{wP} = N_P \times \ell_t = 2790 \times 0.213 = 594\ m$

Wire length of one 4Ω secondary section: $\ell_{w4\Omega} = N_{S4\Omega} \times \ell_t = 94 \times 0.213 = 20\ m$

Wire length of one additional secondary section: $\ell_{wSadd} = N_{Sadd} \times \ell_t = 39 \times 0.213 = 8.3\ m$

DC resistance of the winding ($R_{c\text{-}dc}$):

$$R_{c-DC} = \rho \times \frac{4 \times \ell_w}{\pi \times d^2}$$

ρ – specific resistance of the wire **(Cu wire: $\boldsymbol{\rho = 0.0175\Omega \frac{mm^2}{m}}$)**

DC resistance of the primary winding (total): $\boldsymbol{R_{P-DC}} = \rho \times \frac{4\times\ell_w}{\pi\times d^2} = 0.0175 \times \frac{4\times594}{3.14\times0.265^2} = \mathbf{188.5\ \Omega}$

DC resistance of one 4Ω secondary section: $\boldsymbol{R_{4\Omega-DC}} = \rho \times \frac{4\times\ell_w}{\pi\times d^2} = 0.0175 \times \frac{4\times20}{3.14\times0.9^2} = 0.55\ \Omega$

Since the two secondary 4Ω sections are connected in parallel the total DC resistance of the 4Ω secondary winding is: 0.55 / 2 = 0.275Ω

DC resistance of one additional secondary section: $\boldsymbol{R_{add-DC}} = \rho \times \frac{4\times\ell_w}{\pi\times d^2} = 0.0175 \times \frac{4\times8.3}{3.14\times1.12^2} = 0.147\ \Omega$

Each additional section is connected in series with the secondary 4Ω section so the DC resistance of one secondary 8Ω section is: $R_{4\Omega\text{-}dc} + R_{add\text{-}dc}$ *= 0.55 Ω + 0.147 Ω = 0.697 Ω.*

Since the two secondary 8Ω sections are connected in parallel the total DC resistance of the 8Ω secondary winding is: 0.697 / 2 = 0.348 Ω

<u>Calculation and checking of copper (Cu) losses in transformer windings:</u>

***Cu loss in Primary winding*:** $\boldsymbol{P_{P-loss}}(\%) = 100 \times \frac{R_{P(c-DC)}}{R_{P(c-DC)}+R_{La}} = 100 \times \frac{188.5}{188.5+3500} = \mathbf{5.1}\ \%$

***Cu loss in Secondary winding* (4 Ω):** $\boldsymbol{P_{S4\Omega-loss}}(\%) = 100 \times \frac{R_{S4\Omega(c-dc)}}{R_{S4\Omega(c-DC)}+R_S} = 100 \times \frac{0.275}{0.275+4} = \mathbf{6.4}\ \%$

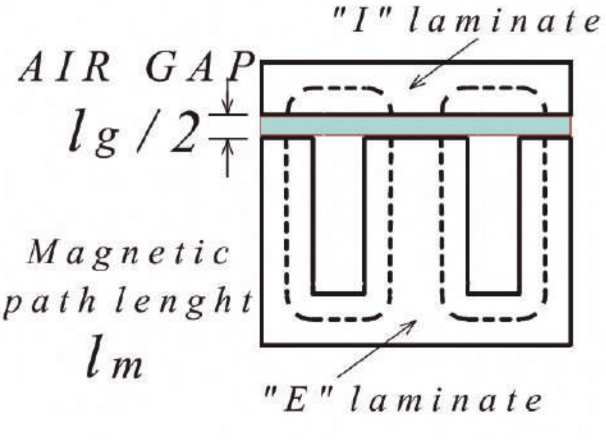
***Total Cu loss (for 4 Ω secondary)*:** $\boldsymbol{P_{loss}}(\%) = P_{P-loss}(\%) + P_{S4\Omega-loss}(\%) = 5.1 + 6.4 = \mathbf{11.5}\ \%$

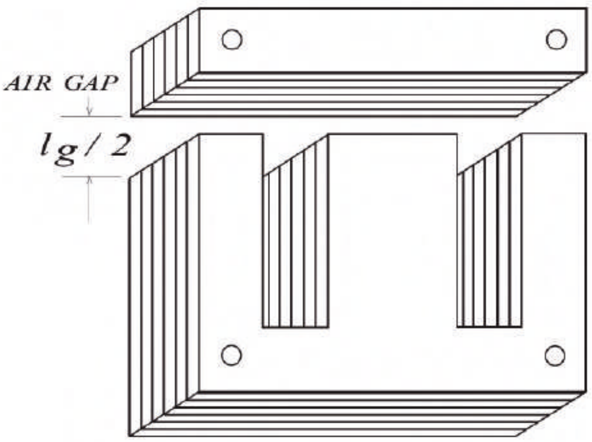
***Cu loss in Secondary winding* (8 Ω):** $\boldsymbol{P_{S8\Omega-loss}}(\%) = 100 \times \frac{R_{S8\Omega(c-DC)}}{R_{S8\Omega(c-DC)}+R_S} = 100 \times \frac{0.348}{0.348+8} = \mathbf{4.1}\ \%$

***Total Cu loss (for 8 Ω secondary)*:** $\boldsymbol{P_{loss}}(\%) = \boldsymbol{P_{P-loss}}(\%) + \boldsymbol{P_{S-loss}}(\%) = 5.1 + 4.1 = \mathbf{9.2}\ \%$

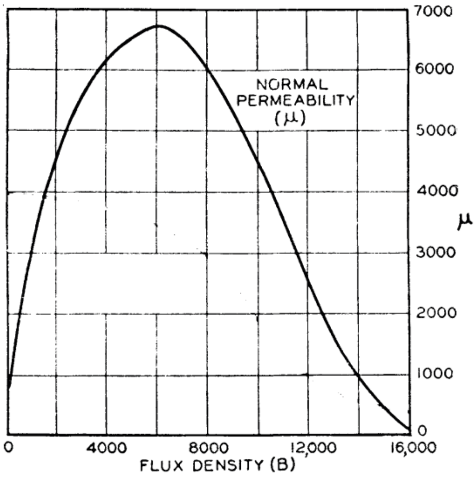
Air gap calculation:
The most important effect of air gap is to prevent the saturation of the transformer core caused by a high DC current flowing through the primary winding or by a high amplitude AC signal.

An air gap is used in the design of single-ended transformers to avoid problems with saturation of the transformer core and to adjust the primary inductance of the transformer to the required value in an easy and simple way.
In practice, the transformer core consists of package of stacked E and a package of stacked I transformer laminate separated by some insulation material (simulating an air gap with a material such as pertinax, mica, paper, plastic foil…) of a defined thickness.





The size of the air gap depends on the material characteristics of the transformer laminates, i.e. the permeability (μ_m) of the transformer laminate and magnetic path length MPL (l_m) of the transformer core.



Equations used in calculating the air gap size:

Calculation of the **effective permeability μ_{eff}** of the transformer core:

$$\mu_{eff} = \frac{B_{DC} \times \ell_m}{1.26 \times N_P \times I_{DC}} \qquad (501.44)$$

SE output transformer: recommended B_{dc} = (4000 ÷ 5000) Gauss

$$\mu_{eff} = \frac{\mu}{1 + \mu \times \frac{\ell_g}{\ell_m}} \qquad (501.45)$$

The size of the air gap ℓ_g :

$$\ell_g = \ell_m \times \frac{(\mu - \mu_{eff})}{\mu \times \mu_{eff}} \qquad (501.46)$$

μ – maximum permeability of the transformer laminate (NOSS transformer laminate: recommended **μ_{max}= 3000)**
μ_{eff} **–** effective permeability
ℓ_m **–** magnetic path length MPL [cm]
ℓ_g **–** air gap [cm]

Very important note: the thickness of the insulation material (δ) placed between the package of stacked E and a package of stacked I transformer laminate is equal to half of the air gap l_g:

$$\delta = \frac{\ell_g}{2} \qquad (501.47)$$

Single ended output transformer: recommended effective permeability μ_{eff} (transformer core: 4% Si transformer laminate):

NOSS: **200 ≤ μ_{eff} ≤ 450 (**recommendation: **250 ≤ μ_e ≤ 350);** [μ = 3000]

GOSS: **200 ≤ μ_{eff} ≤ 750 (**recommendation: **250 ≤ μ_e ≤ 450);** [μ = 5000 ÷ 7000]
If ℓ_g changes, μ_e must be recalculated:

$$\mu_e = \frac{\mu_{max}}{1 + \mu_{max} \times \frac{l_g}{l_{Fe}}}$$

Example (cont.):
EI 108 transformer laminate, Coil former (3.6 x 4.5) cm, $S_{Fe\ eff}$ = 15.55cm^2
N_P = 2790 Turns: 18 layers of 155 turns each. Insulated wire: $d_{P(standard)}$ = 0.265 mm, $d_{Wp(standard)}$ = 0.312 mm
N_S = 94 Turns

Calculating μ_e:

$$\mu_e = \frac{B_{DC} \times l_{Fe}}{1.26 \times N_p \times I_{dc}} = \frac{5000 \times 5.57 \times 3.6}{1.26 \times 2790 \times 0.081} = \mathbf{352}$$

Air gap (ℓ_g):

$$\boldsymbol{l_g} = \frac{l_{Fe} \times (\mu_{max} - \mu_e)}{\mu_{max} \times \mu_e} = \frac{5.57 \times 3.6 \times (3000 - 352)}{3000 \times 352} = 0.05\ cm = \mathbf{0.5\ mm}$$

Gap between E and I laminate stack (δ):

$$\boldsymbol{\delta} = \frac{l_g}{2} = \frac{0.5}{2} = \mathbf{0.25\ mm}$$

Check the inductance of the primary winding (L_P):

$$\boldsymbol{L_P} = \frac{1.256 \times S_{eff} \times \mu_e \times N_p^2}{10^8 \times l_{Fe}}$$

Example (cont.):
EI 108, Coil former (3.6 x 4.5) cm, Effective cross-section of the transformer core $S_{Fe\ eff}$ = 15.55 cm^2

N_P = 2790Turns: 18 layers of 155 turns each. Insulated wire $d_{P(standard)}$ = 0.265 mm = $d_{Wp(standard)}$ = 0.312 mm
N_S = 94 Turns

$$\boldsymbol{L_P} = \frac{1.256 \times S_{eff} \times \mu_e \times N_p^2}{10^8 \times l_{Fe}} = \frac{1.256 \times 15.55 \times 352 \times 2790^2}{10^8 \times 5.57 \times 3.6} = \mathbf{26.7\ H}$$

$$\boldsymbol{R_A} = \frac{(R_i + R_{P-DC}) \times R_{La}}{R_i + R_{P-DC} + R_{La}}$$

$$\boldsymbol{R_A} = \frac{(R_i + R_{P-DC}) \times R_{La}}{R_i + R_{P-DC} + R_{La}} = \frac{(720 + 188.5) \times 3500}{720 + 188.5 + 3500} = \mathbf{721\ \Omega}$$

Calculating *$f_{low\ (-3dB)}$* (for calculated L_P):

$$\mathrm{f_{low}} = \frac{\mathrm{R_A}}{2 \times \pi \times \mathrm{L_P}}$$

$$\mathrm{f_{low}} = \frac{\mathrm{R_A}}{2 \times \pi \times \mathrm{L_P}} = \frac{721}{2 \times 3.14 \times 26.7} = \mathbf{4.3\ Hz}$$

Calculating f_{Co} :

$$\boldsymbol{f_{C_0}} = \frac{R_{La}}{2 \times \pi \times L_p}$$

$$\boldsymbol{f_{C_0}} = \frac{R_{La}}{2 \times \pi \times L_p} = \frac{3500}{2 \times 3.14 \times \times 26.7} = \mathbf{20.8\ Hz}$$

Calculating B_{AC} and B_{DC} :

$$B_{AC} = \frac{U_{eff} \times 10^8}{4.44 \times f_{C_0} \times N_p \times S_{Fe_{eff}}}$$

$$B_{AC} = \frac{U_{eff} \times 10^8}{4.44 \times f_{C_0} \times N_p \times S_{Fe_{eff}}} = \frac{10^8 \times \sqrt{P_{OUT} \times R_{La}}}{4.44 \times f_{C_0} \times N_p \times S_{Fe_{eff}}} = \frac{10^8 \times \sqrt{8.5 \times 3500}}{4.44 \times 20.8 \times 2790 \times 15.55} = \mathbf{4305\ Gauss}$$

$$B_{DC} = \frac{1.26 \times \mu_e \times N_p \times I_{DC}}{l_{Fe}}$$

$$B_{DC} = \frac{1.26 \times \mu_e \times N_p \times I_{DC}}{l_{Fe}} = \frac{1.26 \times 352 \times 2790 \times 0.081}{5.57 \times 3.6} = \mathbf{4998\ Gauss}$$

$$B_{AC} + B_{DC} < B_{max} \ (= \mathbf{13000\ Gauss})$$

$$B_{AC} + B_{DC} < B_{max} \rightarrow (\mathbf{4305 + 4998}) = \mathbf{9303\ Gauss} < (\mathbf{13000\ Gauss})$$

Calculating saturation frequency f_{sat} :

$$f_{sat} = \frac{22.6 \times U_{eff} \times 10^6}{S_{eff} \times N_p \times B_{max}}$$

$$f_{sat} = \frac{22.6 \times U_{eff} \times 10^6}{S_{eff} \times N_p \times B_{max}} = \frac{22.6 \times 10^6 \times \sqrt{8.5 \times 3500}}{15.55 \times 2790 \times 13000} = \mathbf{6.9\ Hz}$$

Leakage inductance and high cut-off frequency

Leakage inductance:
The leakage inductance is mainly determined by the chosen winding geometry.

Example (cont.):
EI 108 transformer laminate, coil former (3.6 x 4.5) cm, Effective cross-section of the core $S_{Fe\ eff}$ = 15.55 cm^2
N_P = 2790Turns:18 layers of 155 turns each.

$$L_S = \frac{0.417 \times Np^2 \times l_t \times [(2 \times n_{P-S} \times \Delta_{ii}) + h_{total}]}{10^9 \times n_{P-d}^2 \times c''}$$

$$L_S = \frac{0.417 \times 2790^2 \times 159 \times [(2 \times 4 \times 0.4) + 13.91]}{10^9 \times 4^2 \times 48.5} = \mathbf{0.011\ H = 11\ mH}$$

$$R_B = R_i + R_{P-DC} + R_{La} = 720\Omega + 188.5\Omega + 3500\Omega = \mathbf{4408.5\ \Omega}$$

$$f_{(-3dB)high} = \frac{R_B}{2\pi \times L_S} = \frac{4360.8}{2 \times 3.14 \times 0.011} = \mathbf{63785\ Hz}$$

Conclusion:
SE output transformer (output tube: WE 300B):
WE 300B: R_i = 720 Ω, I_a = 81 mA, U_a = 350 V
P_{out} = 8.5 W
R_{La} = 3500 Ω

$L_{a(min)}$ = 24.3 H (Minimum inductivity necessary to achieve 95 % full power or – 0.22 dB at 18 Hz)

Transformer laminate: EI 108 F_e S_i (4%) 0.35 mm Transformer coil former: (3.6 x 4.5) cm

Effective cross-section of the transformer core: 15.55 cm^2
Primary: 2790 turns, Cul Ø 0.265 mm

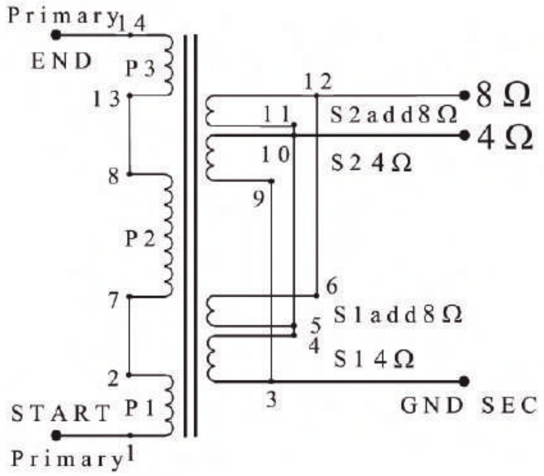
Secondary (4Ω): 94 turns, Cul Ø 0.9 mm (x 2 connected in parallel)
Secondary (8Ω): 133 turns (94 turns, Cul Ø 0.9 mm + 39 turns Cul Ø1.12 mm) x 2 connected in parallel
The complete winding arrangement is:

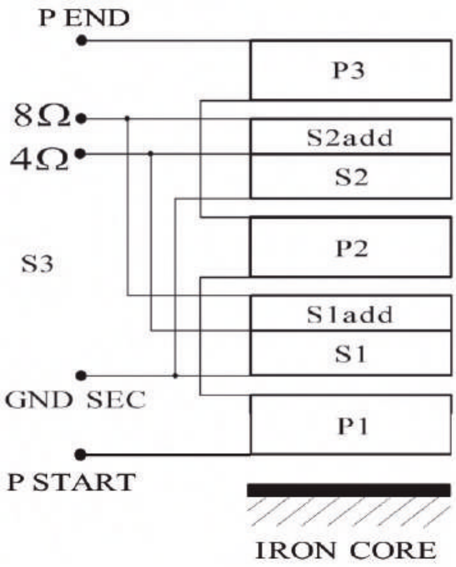
$2S + 3P: \quad (4/18)P - S_{4\Omega}+S_{8\Omega add} - (10/18)P - S_{4\Omega}+S_{8\Omega add} - (4/18)P$

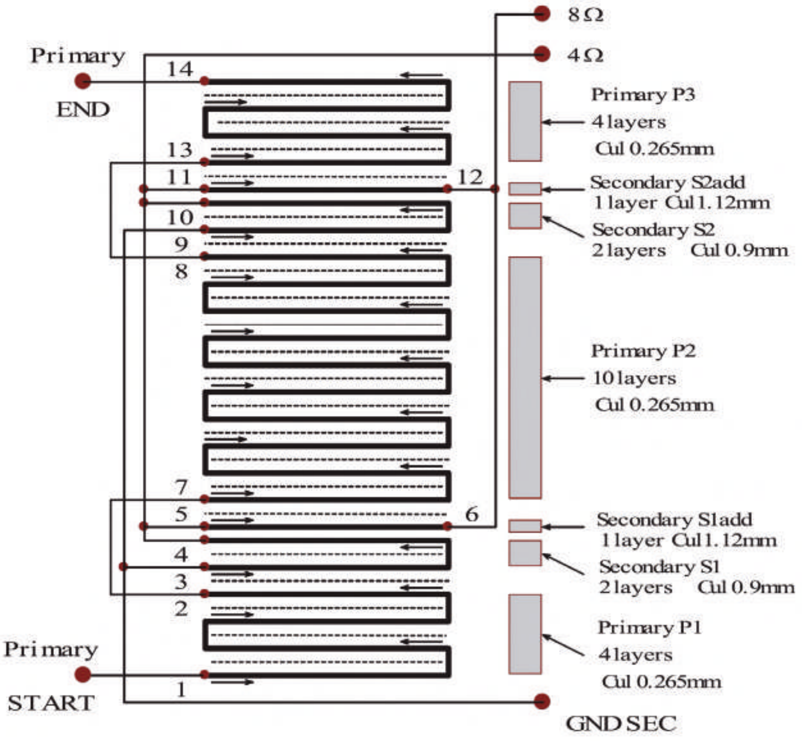
- Primary section P1: 4 layers of155 turns each (total: 620 turns),wire: Cul Ø 0.265 mm, insulation foil between winding layers: 0.05 mm
- Insulation foil: (0.3 ÷ 0.4) mm
- Secondary section $S1_{4\Omega}$: 2 layers of 47 turns each (total: 94 turns), wire: Cul Ø 0.9 mm, insulation foil between winding layers: 0.05 mm
- Insulation foil: 0.05 mm
- Additional secondary section $S1_{add}$: 1 layer of 39 turns, wire: Cul Ø 1.12 mm
- Insulation foil: (0.3 ÷ 0.4) mm
- Primary section P2: 10 layers of 155 turns each (total:1550 turns), wire: Cul Ø 0.265 mm, insulation foil between winding layers: 0.05 mm
- Insulation foil: (0.3 ÷ 0.4) mm
- Secondary section $S2_{4\Omega}$:2 layers of 47 turns each (total: 94 turns), wire: Cul Ø 0.9 mm, insulation foil between winding layers: 0.05 mm
- Insulation foil 0.05mm
- Additional secondary section $S2_{add}$: 1 layer of 39 turns, wire: Cul Ø 1.12 mm
- Insulation foil: (0.3 ÷ 0.4) mm
- Primary section P3: 4 layers of 155 turns each (total: 620 turns), wire: Cul Ø 0.265 mm, insulation foil between winding layers: 0.05 mm

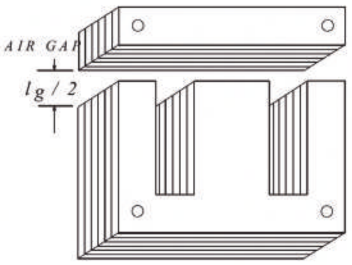
Gap (insulation foil) between E and I laminate stack (δ): 0.25 mm
Effective permeability μ_e *: 352*
Inductance of the Primary winding: 26.7 H
DC resistance of the Primary winding: 188.5 Ω
Cu loss in Primary winding: 5.1%
Cu loss in Secondary (4Ω) winding: 6.4 %
Cu loss in Secondary (8Ω) winding: 4.1 %
Total Cu loss (4Ω Load): 11.5 %
Total Cu loss (8Ω Load): 9.2 %
f_{Co}*: 20.8 Hz*
$f_{low(-3dB)}$*: 4.3 Hz*
$f_{high(-3dB)}$*: 63 kHz*
Leakage inductance: 11mH
DC Magnetic Flux Density: 5000 Gauss
AC Magnetic Flux Density: 4300 Gauss
Total Magnetic Flux Density: 9300 Gauss
Saturation frequency: 6.9 Hz

Transformer design (schematic and graphic representation):

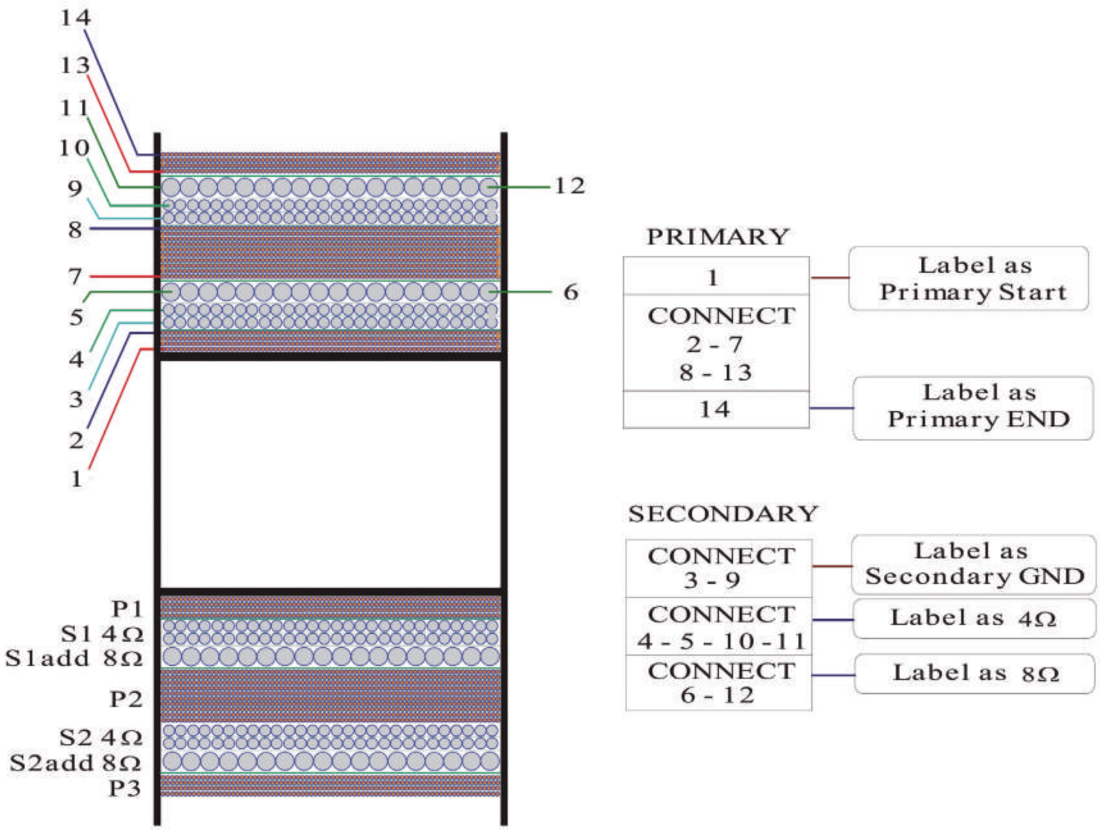








E I 108 (3.6 x 4.5) cm



P1 (1 – 2): 4 layers of 155 turns each (total: 620 turns), wire: Cul Ø 0.265mm; Insulation foil between layers # 0.05mm
S1 (3 – 4): 2 layers of 47 turns each (total 94 turns), wire: Cul Ø 0.9 mm; Insulation foil between layers # 0.05mm
S1add (5 – 6): 1 layer of 39 turns, wire: Cul Ø 1.12 mm
P2 (7 – 8): 10 layers of 155 turns each (total: 1550 turns), wire: Cul Ø 0.265mm; Insulation foil between layers # 0.05mm
S2 (9 – 10): 2 layers of 47 turns each (total 94 turns), wire: Cul Ø 0.9 mm; Insulation foil between layers # 0.05mm
S2add (11 – 12): 1 layer of 39 turns, wire: Cul Ø 1.12 mm
P3 (13 – 14): 4 layers of 155 turns each (total: 620 turns), wire: Cul Ø 0.265mm; Insulation foil between layers # 0.05mm
Insulation foil between P and S sections: # 0.4 mm

 # Note
The DC resistance of the Primary winding determines the Cu loss of the transformer, but also affects the *construction of the high voltage DC power supply stage.*

The DC resistance of the Primary winding is connected in series with the anode of the output tube, so the voltage *drop across the DC resistance of the Primary winding ($I_a \times R_{P\text{-}DC}$) must be taken into account when determining the supply stage voltage (U_{ab}).*

Example above:

a) *Fixed bias :* $U_{ab} = U_a + I_a \times R_{P\text{-}DC} = 350\ V + 0.081\ A \times 188.5\ \Omega = 365.3\ V$

b) *Automatic bias:* $U_{ab} = U_{A\text{-}K} + U_{G\text{-}K}\ (U_K) + I_a \times R_{P\text{-}DC} = 350\ V + 71\ V + 0.081\ A \times 188.5\ \Omega = 436.3\ V$

Example (cont.)
** The above example but using EI 120 transformer laminate and coil former: (4 x 4) cm. The same winding geometry is applied.*

Designing a SE output transformer (output tube: WE 300B):

WE 300B: $R_i = 720\ \Omega$, $I_a = 81\ mA$, $U_a = 350\ V$
$P_{out} = 8.5\ W$
$R_{La} = 3500\ \Omega$
$L_{a(min)} = 24.3\ H$ *(95% of full power or power attenuation of – 0.22 dB at 18 Hz)*

Transformer laminate: EI 120, $F_e\ S_i$ *(4%) 0.35 mm*
Transformer coil former: (4 x 4) cm

Effective cross-section of the transformer core: 15.36 cm^2
Primary: 2898 turns, wire: Cul Ø 0.28 mm
Secondary (4Ω): 98 turns Cul Ø 1 mm (x 2 connected in parallel)
Secondary (8Ω): 138 turns (98 turns Cul Ø 1 mm + 40 turns Cul Ø 1.18 mm) x 2 connected in parallel

Winding arrangement:

2S + 3P: $(4/18)P - S_{4\Omega}+S_{8\Omega add} - (10/18)P - S_{4\Omega}+S_{8\Omega add} - (4/18)P$

- *P1: 4 layers of 161 turns each (total: 644 turns), wire: Cul Ø 0.28 mm; Insulation foil between layers # 0.05 mm*
- *Insulation foil: (0.3 ÷ 0.4) mm*
- $S1_{4\Omega}$ *: 2 layers of 49 turns each (total: 98 turns), wire: Cul Ø 1mm; Insulation foil between layers # 0.05 mm*
- *Insulation foil: 0.05 mm*
- $S1_{add}$ *: 1 layer of 40 turns, wire: Cul Ø 1.18 mm*
- *Insulation foil: (0.3 ÷ 0.4) mm*
- *P2: 10 layers of 161 turns each (total: 1610 turns), wire: Cul Ø 0.28 mm; Insulation foil between layers # 0.05 mm*
- *Insulation foil: (0.3 ÷ 0.4) mm*
- $S2_{4\Omega}$ *:2 layers of 49* turns each (total: 98 turns), wire: Cul Ø 1mm; Insulation foil between layers # 0.05 mm
- *Insulation foil: 0.05 mm*
- $S2_{add}$ *: 1 layer of 40 turns , wire: Cul Ø 1.18 mm*
- *Insulation foil: (0.3 to 0.4) mm*
- *P3: 4 layers of 161 turns each (total: 644 turns), wire: Cul Ø 0.28mm; Insulation foil between layers # 0.0 5 mm*

Gap (insulation foil) between E and I laminate stack (δ): 0.25 mm
Effective permeability μ_e *: 376*
Inductance of the Primary winding: 27.4 H
DC resistance of the Primary winding: 177 Ω
Cu loss in Primary winding: 4.8 %
Cu loss in Secondary (4Ω) winding: 5.5 %
Cu loss in Secondary (8Ω) winding: 3.6 %
Total Cu loss (4Ω Load): 10.3 %
Total Cu loss (8Ω Load): 8.4 %
f_{Co}*: 20.35 Hz*
$f_{low(-3dB)}$*: 4Hz*
$f_{high(-3dB)}$*: 51.3 kHz*
Leakage inductance: 13.5 mH
DC Magnetic Flux Density: 5000 Gauss
AC Magnetic Flux Density: 4288 Gauss
Total Magnetic Flux Density: 9288 Gauss
Saturation frequency: 9.4Hz

With the skill and experience of the designer as well as the use of ***engineering tricks****, the characteristics of the transformer can improve (for example: reduce the leakage inductance and improve the characteristics of the transformer at high frequencies) or the production of the transformer can be made easier or simplified.*

The Single Ended Output transformer 3500Ω / 4Ω / 8Ω, 8.5W described in the above example (EI 108 transformer laminate and coil former (3.6 x 4.5) cm) can be constructed in another way:
The winding arrangement can be applied: **4P + 3S** *: (P/6 – S – P/3 – S – P/3 – S – P/6).*

Primary sections connected in series, Secondary sections connected in parallel.

First trick:
One section of 4 Ω Secondary winding (94 turns) can be wound as follows:

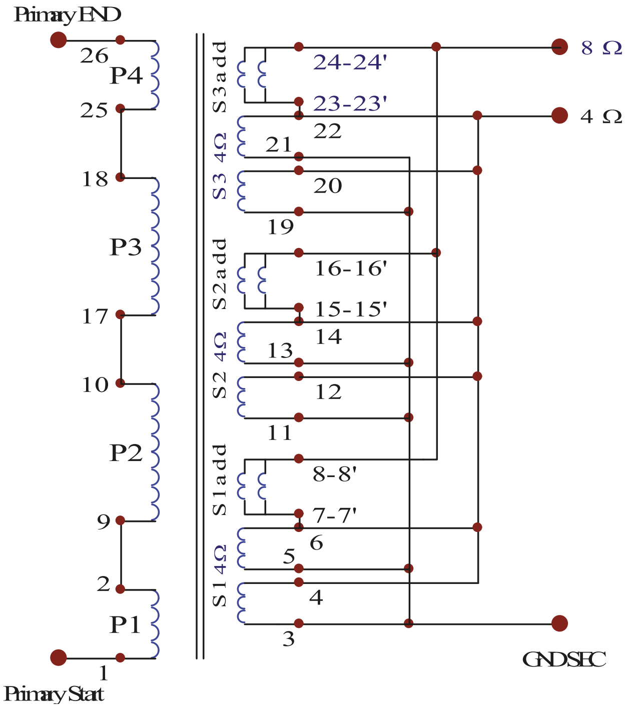
- *EI 108 coil former: 94 turns of insulated wire max. diameter* $d_{Wp} \leq [0.97 \times (c - 2s)] / N_S = [0.97 \times (54 - 4)\ \text{mm}] / 94 = 0.515$ *mm. can be wound in one layer; Standard Cu wire Ø 0.45 mm* ($d_{Wp} = 0.512$ *mm) can be used.*
- ***Each secondary section is made of two parallel connected windings each of 94 turns wound in one layer with CuI Ø 0.45 mm wire****.*
- *Each secondary section can handle a current of 0.4757 A.*
- As the three secondary sections are connected in parallel, the Secondary winding can handle a current of:
- $I_{S\ 4\Omega} = 3 \times 0.4757\ \text{A} = 1.427\ \text{A}$ ($J = 1.5\ \text{A} / \text{mm}^2$) or, $I_{S\ 4\Omega} = 1.9\ \text{A}$ ($J = 2\ \text{A} / \text{mm}^2$).

Second trick:
The 8 Ω Secondary winding section is made by winding an additional winding (above example (EI 108): 1 layer of 39 turns of standard CuI Ø 1.12 mm wire) and connecting it in series with the 4 Ω Secondary winding section.

The additional winding can be wound using a special winding method – winding with ***two wires*** *or so-called* ***bifilar*** *winding. In fact, a bifilar winding consists of two closely spaced, parallel windings that can be connected in different ways: a) parallel-winding, series connection, b) parallel-winding, parallel connection, c) counter-winding (series), d) counter-winding (parallel). Therefore, the additional secondary sections in this example can be wound using type b) parallel-winding, parallel connection bifilar winding method: one layer of 39 turns of two CuI Ø* 0.53mm *wires (insulated wire: Ø 601mm). The benefit of this method of winding the additional winding consists in reducing the height of the winding.*

Winding arrangement: **4P + 3S** : (P/6 – S – P/3 – S – P/3 – S – P/6).

- P1: 3 *layers of 155 turns each (total: 465 turns), wire: CuI Ø 0.265mm; Insulation foil between layers # 0.05 mm*
- *Insulation foil: (0.3 ... 0.4) mm*
- $S1_{4\Omega}$ *: 2 windings of 94 turns each of CuI Ø 0.45mm connected in parallel. Each winding is wound in one layer. Insulation foil between layers # 0.05 mm*
- *Insulation foil # 0.05mm*
- $S1_{add}$ *:1 layer of 39 turns of two CuI Ø 0.53mm (bifilar: parallel-winding, parallel connection)*
- *Insulation foil: (0.3 ... 0.4) mm*
- *P2: 6 layers of 155 turns each (total: 930 turns), wire: CuI Ø 0.265mm; Insulation foil between layers # 0.05 mm*
- *Insulation foil: (0.3 ... 0.4) mm*
- $S2_{4\Omega}$ *: 2 windings of 94 turns each of CuI Ø 0.45mm connected in parallel. Each winding is wound in one layer. insulation foil between layers # 0.05 mm*
- *Insulation foil # 0.05 mm*
- $S2_{add}$ *: 1 layer of 39 turns of two CuI Ø 0.53mm (bifilar: parallel-winding, parallel connection)*
- *Insulation foil: (0.3 ... 0.4) mm*
- *P3: 6 layers of 155 turns each (total: 930 turns), wire: CuI Ø 0.265mm; Insulation foil between layers # 0.05 mm*
- *Insulation foil: (0.3 ... 0.4) mm*
- $S3_{4\Omega}$ *: 2 windings of 94 turns each of CuI Ø 0.45mm connected in parallel. Each winding is wound in one layer. insulation foil between layers # 0.05 mm*
- *Insulation foil # 0.05 mm*
- $S3_{add}$ *: 1 layer of 39 turns of two CuI Ø 0.53mm (bifilar: parallel-winding, parallel connection)*
- *Insulation foil: (0.3 ... 0.4) mm*
- *P4: 3 layers of 155 turns each (total: 465 turns), wire: CuI Ø 0.265mm; Insulation foil between layers # 0.05 mm*

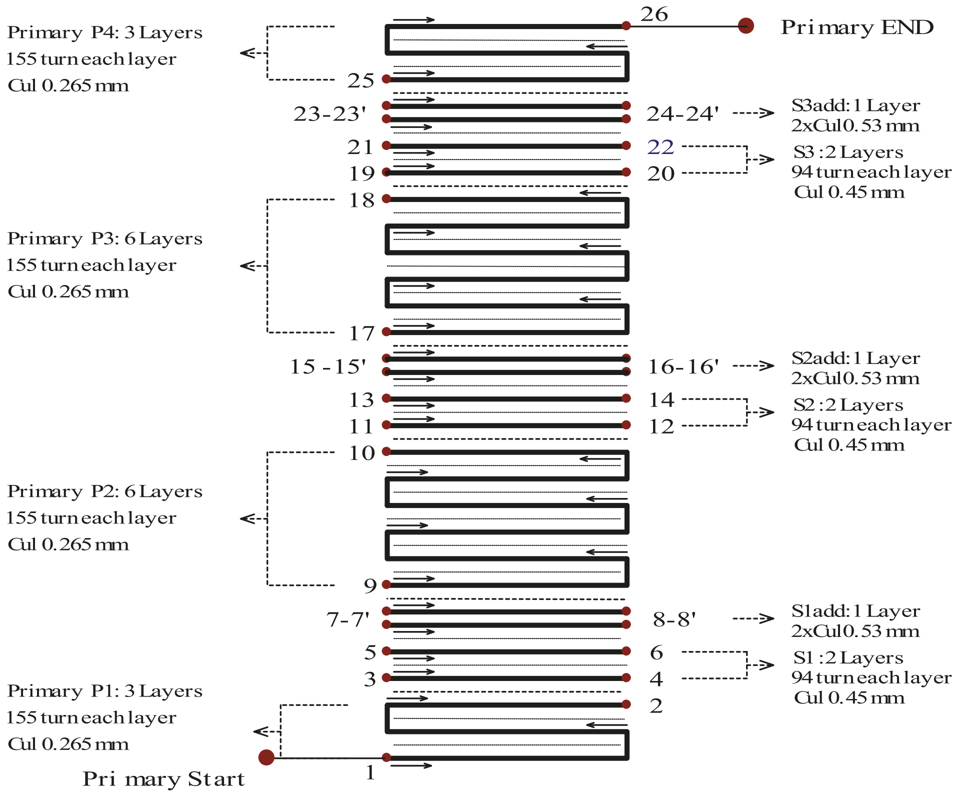


PRIMARY

1	START
CONNECT 2-9 10-17 18-25	
26	END

SECONDARY

CONNECT 3-5-11-13-19-21	GND
CONNECT 4-6-7-7'-12-14-15-15'-20-22-23-23'	4Ω
CONNECT 8-8'-16-16'-24-24'	8Ω



The leakage inductance is:

$$L_S = \frac{0.417 \times 2790^2 \times 159 \times [(2 \times 4 \times 0.4) + 13.91]}{10^9 \times 6^2 \times 48.5} = \mathbf{0.005\ H = 5\ \mathit{mH}}$$

$$f_{(-3dB)high} = \frac{R_B}{2\pi \times L_S} = \frac{4360.8}{2 \times 3.14 \times 0.005} = \mathbf{138879\ \mathit{Hz} = 138.9\ \mathit{kHz}}$$

** The new arrangement of the windings as well as the use of the winding methods described above improve the characteristics of the transformer at high frequencies:*

$f_{(-3dB)\ high}$ = 138.9 kHz. (previous winding arrangement: $f_{(-3dB)\ high}$ = 63 kHz).

Output transformer design flow chart:

FIRST STEP

PROJECT REQUIREMENTS – INPUT DATA:

- The operating point of the output tube (Quiescent point).
- Internal resistance (R_i) of the output tube
- Plate (anode) resistance – plate (anode) load (R_{La})
- Output power
- Frequency band
- Secondary load

SECOND STEP

- Calculating the cross- section of the iron core

$$S_{Fe}(cm^2) = \frac{1.8\times10^6\times P_p(W)}{f(Hz)\times B(gauss)\times S_W(cm^2)\times J(\frac{A}{mm^2})} \qquad S_{Fe}(cm^2) \approx \frac{(14\div17)\times P(W)}{S_W(cm^2)}$$

$$S_{Fe(min)}(cm^2) = \sqrt{\frac{P_{0maxac}\times G\times10^6}{B\times J\times f}} \qquad S_{Fe}(cm^2) \approx (4.3 \div 4.9)\times\sqrt{P_{0_{max}}(W)}$$

- Choosing a standard transformer laminate and coil former
- Calculating: S_{eff} , S_w ,ℓ_{fe} , V_{Fe} , G_{Fe}
- $S_{Fe\ eff}$ = (0.92 – 0.96) x (b x h)
- S_W = a x c
- ℓ_{Fe} = 2 x c + 2 x a + π x a ℓ_{Fe} = 5.57 x b
- V_{Fe} = $S_{Fe\ eff}$ x / ℓ_{Fe}
- G_{Fe} = D_{Fe} x V_{Fe}

THIRD STEP

- Calculating the number of turns of the Primary and Secondary windings (N_P and N_S)

$$N_p = \frac{10^8\times\sqrt{P\times R_{La}}}{4.44\times f_{min}\times B_{max}\times S_{Fe_{eff}}} \qquad N_P \approx \frac{(200\div300)\times\sqrt{P\times R_{La}}}{S_{Fe_{eff}}}$$

$$\mathrm{N_P} = 12.5\times \mathrm{S_w}\times\frac{\mathrm{J}}{\mathrm{I}} \qquad N_P \approx (20\ \div 25)\times\frac{S_W}{I}$$

$$N_P = \frac{B_{dc}\times l_{Fe}}{1.257\times\ \mu_e\times I_{DC}}$$

$$N_S = N_P\times\sqrt{\frac{R_S}{R_{La}}}$$

FOURTH STEP

- Calculating the minimum diameters of the Primary and Secondary windings wires

$$d_{Pmin}(mm) = 1.13 \times \sqrt{\frac{I_a(A)}{J(\frac{A}{mm^2})}} \qquad I_S = \frac{N_P}{N_S} \times \sqrt{\frac{P}{R_{La}}}$$

$$d_{Wp}(mm) = \sqrt{\frac{0.28 \times S_W(mm^2)}{N_P}} \quad \rightarrow \quad N_{P(max)} = \frac{0.28 \times S_W}{d_{Wp}^2}$$

$$Turn/_{Layer} = \frac{0.97 \times c''}{\mathrm{d_{Wp(standard)}}} = \frac{0.97 \times (c - 4mm)}{\mathrm{d_{Wp(standard)}}\ (mm)} \qquad n_{layer} = \frac{N_P}{\left(\frac{Turns}{Layer}\right)}$$

FIFTH STEP

- Choosing the winding arrangement

SIXTH STEP

- Recalculating the diameter of the wires and checking the possibility of placing all the windings inside the window of the transformer core

$$height_{total} = n_{Players} \times (d_{P\,with\,isolation} + \Delta_i) + n_{Slayers} \times (d_{S\,with\,isolation} + \Delta_i) + n_{P-S\,layers} \times \Delta_{ii}$$

- Recalculating the number of turns of the Primary and Secondary windings

SEVENTH STEP

- Calculating the air gap and δ
- Calculating and checking μ

$$\mu_{eff} = \frac{B_{DC} \times l_{Fe}}{1.26 \times N_p \times I_{DC}} \qquad \mu_{eff} = \frac{\mu}{1 + \mu \times \frac{\ell_g}{\ell_m}}$$

$$\ell_g = \ell_m \times \frac{(\mu - \mu_{eff})}{\mu \times \mu_{eff}} \qquad \delta = \frac{\ell_g}{2}$$

EIGHTH STEP

- Calculating the resistance of the windings
- Calculating the Cu loss

$$\ell_t = 2 \times h + \pi \times h_0 + 2 \times b + 4 \times s \qquad \ell_{wP} = N_P \times \ell_t$$

$$R_{c-DC} = \rho \times \frac{4 \times \ell_w}{\pi \times d^2} \qquad P_{P-loss}(\%) = 100 \times \frac{R_{P(c-DC)}}{R_{P(c-DC)} + R_{La}}$$

$$\boldsymbol{P_{S-loss}}(\%) = 100 \times \frac{R_{S(c-DC)}}{R_{S(c-DC)} + R_S} \qquad P_{loss}(\%) = P_{P-loss}(\%) + P_{S-loss}(\%)$$

NINTH STEP

- Calculating: L_p , L_s, f_{Low}, f_{-3dB}, f_{C0}, B_{DC}, B_{AC}, B_{tot}, f_{sat}

$$L_P = \frac{1.256 \times S_{eff} \times \mu_e \times N_p^2}{10^8 \times l_{Fe}} \qquad R_A = \frac{(R_i + R_{P-DC}) \times R_{La}}{R_i + R_{P-DC} + R_{La}}$$

$$f_{low}(f_{(-3dB)low}) = \frac{R_A}{2 \times \pi \times L_P} \qquad f_{C_0} = \frac{R_{La}}{2 \times \pi \times L_p}$$

$$B_{AC} = \frac{U_{eff} \times 10^8}{4.44 \times f_{C_0} \times N_p \times S_{Fe_{eff}}} \qquad B_{DC} = \frac{1.26 \times \mu_e \times N_p \times I_{dc}}{l_{Fe}}$$

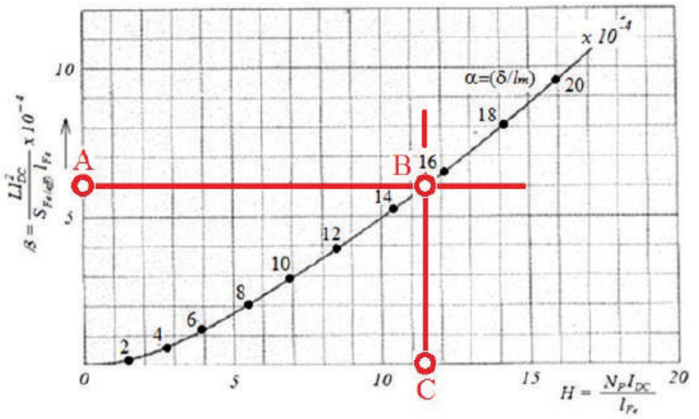
$$f_{sat} = \frac{22.6 \times U_{eff} \times 10^6}{S_{eff} \times N_p \times B_{max}} \qquad L_S = \frac{0.417 \times Np^2 \times l_t \times [(2 \times n_{P-S} \times \Delta_{il}) + h_{total}]}{10^9 \times n_{P-D}^2 \times c''}$$

$$\boldsymbol{R_B} = R_i + R_{P-DC} + R_{La} \qquad f_{(-3dB)high} = \frac{R_B}{2\pi \times L_S}$$

TENTH STEP

- Making, testing and measuring transformer characteristics

A simple method of calculating an SE-Output transformer:



Modified Hanna´s Diagram – metric units

** The original Hanna's diagram uses Imperial units (inches).*

A procedure of calculating SE output transformer using Hanna's diagram

Calculating the cross-section of the iron core S_{Fe} and $S_{Fe\,(eff)}$ (cm^2) and choosing the standard EI laminate is performed according to the procedure explained in the text above (Calculation of the Iron Core Cross - Section (S_{Fe})).

- In order to apply Hanna's method, it is necessary to also calculate:
  - Inductance L_P (H) of the Primary winding.
  - *Magnetic circuit length MPL* = $\boldsymbol{\ell_{Fe}}$ (cm)
  - $\boldsymbol{\beta} = \frac{L_P(H) \times I_{DC}^2(A^2)}{S_{Fe(eff)}(cm^2) \times l_{Fe}(cm)}$
  - *The horizontal line drawn from the point **A** = (the numerical value of β calculated above) located on the ordinate of Hanna's diagram intersects the curve α = (δ / ℓ_m) at point **B**.*
  - *A vertical line drawn from a point **B** intersects abscissa* $H = \left(\frac{N_P \times I_{DC}}{\ell_{Fe}}\right)$ *at point **C**.*
  - Using the numerical value of H at point C, the number of turns of the Primary winding (N_P) can be calculated:

$$\boldsymbol{N_P} = \frac{H \times l_{Fe}(cm)}{I_{DC}\,(A)} = \frac{(numerical\ value\ of\ \boldsymbol{H}\ at\ point\ \boldsymbol{C}) \times l_{Fe}(cm)}{I_{DC}\,(A)}$$

- *The gap between E and I laminate stacks* (δ) can be calculated using the numerical value of the curve *α at point B:*

$$\delta(mm) = 10 \times \alpha \times l_m(cm) = 10 \times (numerical\ value\ of\ the\ curve\ \boldsymbol{\alpha}\ at\ point\ \mathbf{B}) \times l_m(cm)$$

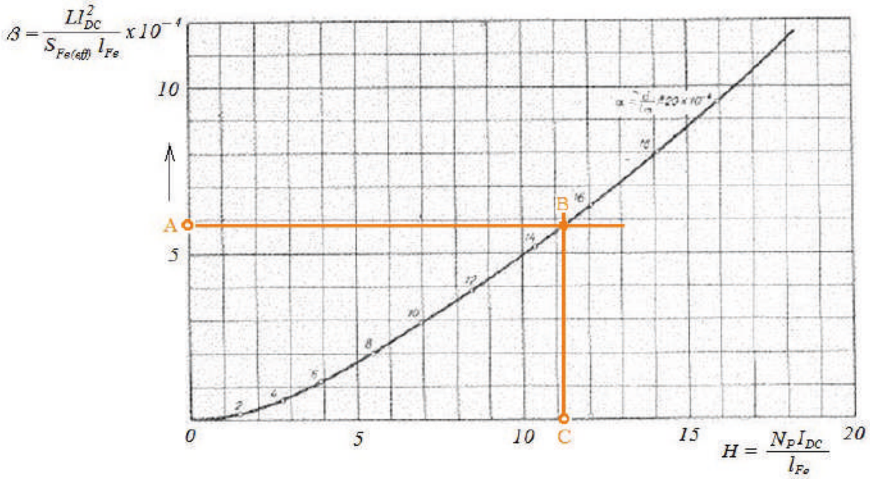
- Designing of the output transformer continues according to the previously explained procedure.

Example
Designing an SE output transformer (output tube: WE 300B):
WE 300B: $R_i = 720\ \Omega$, $I_a = 81$ mA, $U_a = 350$ V, $U_{G\text{-}K} = 71$ V
$P_{out} = 8.5$ W
$R_{La} = 3500\ \Omega$
$L_{a(min)} = 24.3$ H *(95% of full power or power attenuation of – 0.22 dB at 18 Hz)*

Transformer laminate: EI 108 F_e S_i *(4%) 0.35 mm*
Transformer coil former: (3.6 x 4.5) cm
Effective cross-section of the iron core: 15.55 cm^2
$\ell_m = \ell_{Fe} = 20.052$ cm

Calculation of β (point A):

$$\boldsymbol{\beta} = \frac{\boldsymbol{L_P}(H)\times \boldsymbol{I_{DC}^2}(A^2)}{\boldsymbol{S_{Fe(eff)}}(cm^2)\times \boldsymbol{l_{Fe}}(cm)} = \frac{27.85\times 0.081^2}{15.55\times 20.052} = 5.86\times 10^{-4}$$



From the diagram:
Point C: H = 11.2
Point B: $\alpha = 15 \times 10^{-4}$

Calculating the number of turns of the Primary winding (N_P)*:*

$$N_P = \frac{H \times l_{Fe}(cm)}{l_{DC}(A)} = \frac{11.2 \times 20.052}{0.081} = \mathbf{2773}\ Turns$$

The number of turns of the primary winding calculated using the Hanna's method $N_P = 2773$ *is approximately equal to the previously calculated number of turns of the primary winding* $N_P = 2790$.

Calculating the gap between the E and I laminate stacks:

$$\delta(mm) = 10 \times \alpha \times l_m(cm) = 10 \times 15 \times 10^{-4} \times 20.052 = \mathbf{0.3}\ mm$$

*The gap δ = 0.3 mm differs from the previously calculated (δ = 0.25 mm). [*This is due to the different characteristics of the transformer core material used in the previous calculation and the transformer core material used in designing Hanna's diagram].*

Hanna's graphical method is applied to quickly calculate the approximate number of turns of the Primary winding and the air gap of the SE output transformer.

SE OUTPUT TRANSFORMER R_P = 1800Ω 25W

Single Ended output stage: two 300B connected in parallel

Two 300B connected in parallel: Ua = 350 V; Ia = 162 mA (2 x 81 mA); Automatic grid bias: U_{g-c} = - 71V;
Ri = 362.5 Ω (725 Ω / 2);
R_{La} = 1800 Ω; Load 4 Ω, 8 Ω; P_{max} = 25 W;

EI 150N, 4 % Si, Δ = 0.35 mm;
Coil Former: (50 x 50) mm
N_p = 2336 turns
N_{s4ohm} = 110 turns
N_{s8ohm} = 155 turns

Winding arrangement: **5P + 4S** (5 Primary sections – 4 Secondary sections)
(P1) – (S1a, S1b, S1") – (P2) – (S2a, S2b, S2") – (P3) – (S3a, S3b, S3") – (P4) – (S4a, S4b, S4") – (P5)
The total number of layers of the Primary winding: **16**
(2/16) P – $S_{4\Omega}$ II $S_{4\Omega}$+S_{add} – (4/16) P – $S_{4\Omega}$ II $S_{4\Omega}$+S_{add} – (4/16) P – $S_{4\Omega}$ II $S_{4\Omega}$+S_{add} – (4/16) P – $S_{4\Omega}$ – $S_{4\Omega}$+S_{add} – (2/16) P

Primary

Each section of the Primary consists of a certain number of layers of 146 turns of wire Cul Ø 0.4 mm (Ø 0.45 mm with insulation).
P1: 2 layers (292 turns), P2: 4 layers (584 turns), P3: 4 layers (584 turns), P4: 4 layers (584 turns), P5: 2 layers (292 turns)
The Primary sections are connected in series (P1 + P2 + P3 + P4 +P5)

Secondary

4 Ω:

Each section of the 4 Ω Secondary consists of two windings connected in parallel. Each of the windings consists of one layer of 110 turns of wire Cul Ø 0.55 mm (0.6mm with insulation). The sections of the 4Ω Secondary are connected in parallel.
($S_{4\Omega}$ II $S_{4\Omega}$) II ($S_{4\Omega}$ II $S_{4\Omega}$) II ($S_{4\Omega}$ II $S_{4\Omega}$) II ($S_{4\Omega}$ II $S_{4\Omega}$) ; (S1a II S1b) II (S2a II S2b) II (S3a II S3b) II (S4a II S4b)
(110 turns II 110 turns) II (110 turns II 110 turns) II (110 turns II 110 turns) II (110 turns II 110 turns)

8 Ω:

Each section of the additional Secondary winding consists of one layer of 45 turns of two wires Cul Ø 0.7 mm (0.75 mm with insulation) *(bifilar: parallel-winding, parallel connection).* Each section of the additional Secondary winding is connected in series with adjacent section of the 4Ω Secondary making 8 Ω Secondary sections. The sections of the 8Ω Secondary are connected in parallel.
($S_{4\Omega}$II$S_{4\Omega}$ + 45 turns bifilar) II ($S_{4\Omega}$II$S_{4\Omega}$ + 45 turns bifilar) II ($S_{4\Omega}$II$S_{4\Omega}$ + 45 turns bifilar) II ($S_{4\Omega}$II$S_{4\Omega}$ + 45 turns bifilar)

Insulation foil between P layers: 0.05 mm; Insulation foil between P and S layers: 0.4 mm;
Insulation foil between S layers: 0.05 mm

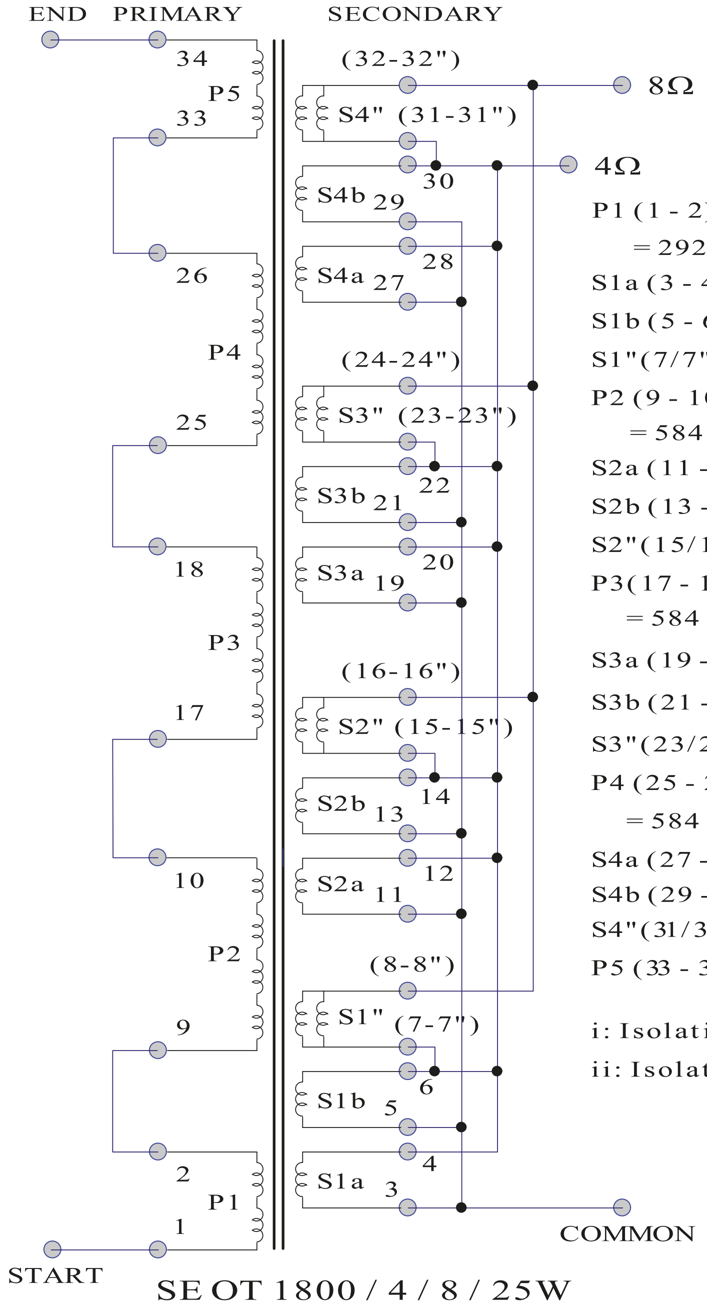
Air gap (production): δ = (0.4 mm)

μ_e = 300
L_P = 17.7 H
L_S = 3.39 mH
R_{P-DC} = 86.48 Ω
f_{C0} = 16.46 Hz
f_{sat} = 6.58 Hz
f_{-3dB} = 3.2 Hz
B_{DC} = 0.5136 T
B_{AC} = 0.5177 T

$\mathbf{U_{ab[\text{auto bias}]}} = U_k + U_{ak} + R_{P\text{-}DC} \times I_a = 71V + 350V + (86.48 \times 0.162A) = \mathbf{435\ V}$

All **E** laminates stack in the same direction and all **I** laminates stack in a same direction.

 # Air gap between E and I laminate stacks: insulation foil # 0.4mm.

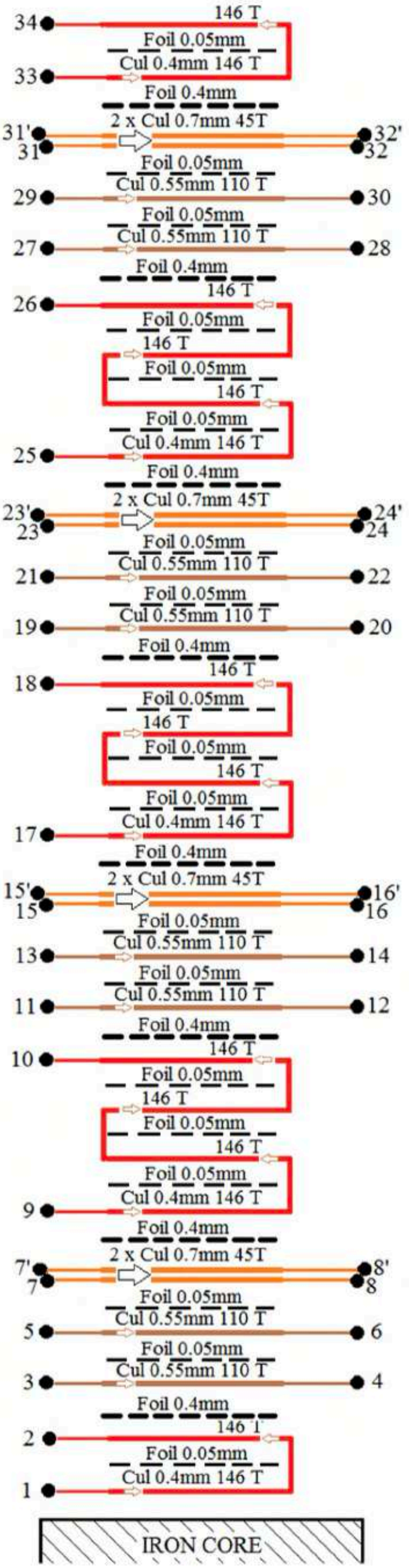


P1 (1 - 2): (146) - i - (146) - ii
= 292Turns Cul 0.4mm
S1a (3 - 4): (110) - i Turns Cul 0.55mm
S1b (5 - 6): (110) - i Turns Cul 0.55mm
S1"(7/7" - 8/8"): (45) - ii Turns 2 x Cul 0.7mm
P2 (9 - 10): (146) - i - (146) - i - (146) - i - (146) - ii
= 584 Turns Cul 0.4mm
S2a (11 - 12): (110) - i Turns Cul 0.55mm
S2b (13 - 14): (110) - i Turns Cul 0.55mm
S2"(15/15" - 16/16"): (45) - ii Turns 2 x Cul 0.7mm
P3(17 - 18): (146) - i - (146) - i - (146) - i - (146) - ii
= 584 Turns Cul 0.4mm
S3a (19 - 20): (110) - i Turns Cul 0.55mm
S3b (21 - 22): (110) - i Turns Cul 0.55mm
S3"(23/23" - 24/24"): (45) - ii Turns 2 x Cul 0.7mm
P4 (25 - 26): (146) - i - (146) - i - (146) - i - (146) - ii
= 584 Turns Cul 0.4mm
S4a (27 - 28): (110) - i Turns Cul 0.55mm
S4b (29 - 30): (110) - i Turns Cul 0.55mm
S4"(31/31" - 32/32"): (45) - ii Turns 2 x Cul 0.7mm
P5 (33 - 34): (146) - i - (146) - ii = 292 Turns Cul 0.4m

i: Isolation Foil 0.05mm
ii: Isolation Foil 0.4mm

Iron Core: EI 150N (Fe 4%Si)
Coil Former: 50mm x 50mm
Air Gap (between E and I): 0.4 mm

SE OT 1800 / 4 / 8 / 25W



PRIMARY

START	CONNECTED	END
1	2 - 9 10 - 17 18 - 25 26 - 33	34

SECONDARY

COMMON

CONNECTED

3 - 5 - 11 - 13 - 19 - 21 - 27 - 29

4 Ω

CONNECTED

4 - 6 - (7-7') - 12 -14 - (15-15') - 20 -
- 22 - (23-23') - 28 - 30 - (31-31')

8 Ω

CONNECTED

(8-8') - (16-16') - (24-24') - (32-32')

SE OT 1800 / 4 / 8 / 25W

SE OUTPUT TRANSFORMER R_P = 2500Ω 25W

Single Ended output stage: two KT88 connected in parallel (Triode or Ultra linear)

R_{La} = 2500Ω; Load 4Ω, 8Ω; P_{max} = 25W; I_{DC} = 180 mA

EI 150N, 4%Si, Δ = 0.35mm;
Coil Former: (50 x 50) mm

N_p = 2520 turns
N_{s4ohm} = 100 turns
N_{s8ohm} = 141 turns

Winding arrangement: **5P + 4S** (5 Primary sections – 4 Secondary sections)
(P1) – (S1, S1add) – (P2) – (S2, S2add) – (P3) – (S3, S3add) – (P4) – (S4, S4add) – (P5)
The total number of layers of the Primary winding: **18**
(2/18) P – $S_{4\Omega}$ II $S_{4\Omega}$+S_{add} – (5/18) P – $S_{4\Omega}$ II $S_{4\Omega}$+S_{add} – (4/18) P – $S_{4\Omega}$ II $S_{4\Omega}$+S_{add} – (5/18) P – $S_{4\Omega}$ – $S_{4\Omega}$+S_{add} – (2/18) P

Primary
Each section of the Primary consists of a certain number of layers of 140 turns of wire Cul Ø 0.425 mm (Ø 0.489 mm with insulation).
P1: 2 layers (280 turns), P2: 5 layers (700 turns), P3: 4 layers (560 turns: 140 turns – UL TAP – 280 turns – UL TAP – 140 turns), P4: 5 layers (700 turns), P5: 2 layers (280 turns)
The Primary sections are connected in series (P1 + P2 + P3 + P4 +P5)

Secondary
4 Ω:
Each section of the 4 Ω Secondary consists of two windings connected in parallel. Each of the windings consists of one layer of 100 turns of wire Cul Ø 0.6 mm (0.675 mm with insulation). The sections of the 4Ω Secondary are connected in parallel.
($S_{4\Omega}$ II $S_{4\Omega}$) II ($S_{4\Omega}$ II $S_{4\Omega}$) II ($S_{4\Omega}$ II $S_{4\Omega}$) II ($S_{4\Omega}$ II $S_{4\Omega}$) ; (S1a II S1b) II (S2a II S2b) II (S3a II S3b) II (S4a II S4b)
(100 turns II 100 turns) II (100 turns II 100 turns) II (100 turns II 100 turns) II (100 turns II 100 turns)

8 Ω:
Each section of the additional Secondary winding consists of one layer of 41 turns of two wires Cul Ø 0.75 mm (0.832 mm with insulation) *(bifilar: parallel-winding, parallel connection).* Each section of the additional Secondary winding is connected in series with adjacent section of the 4Ω Secondary making 8 Ω Secondary sections. The sections of the 8Ω Secondary are connected in parallel.
($S_{4\Omega}$II$S_{4\Omega}$ + 41 turns bifilar) II ($S_{4\Omega}$II$S_{4\Omega}$ + 41 turns bifilar) II ($S_{4\Omega}$II$S_{4\Omega}$ + 41 turns bifilar) II ($S_{4\Omega}$II$S_{4\Omega}$ + 41 turns bifilar)

Insulation foil between P layers: 0.05 mm; Insulation foil between P and S layers: 0.4 mm;
Insulation foil between S layers: 0.05 mm

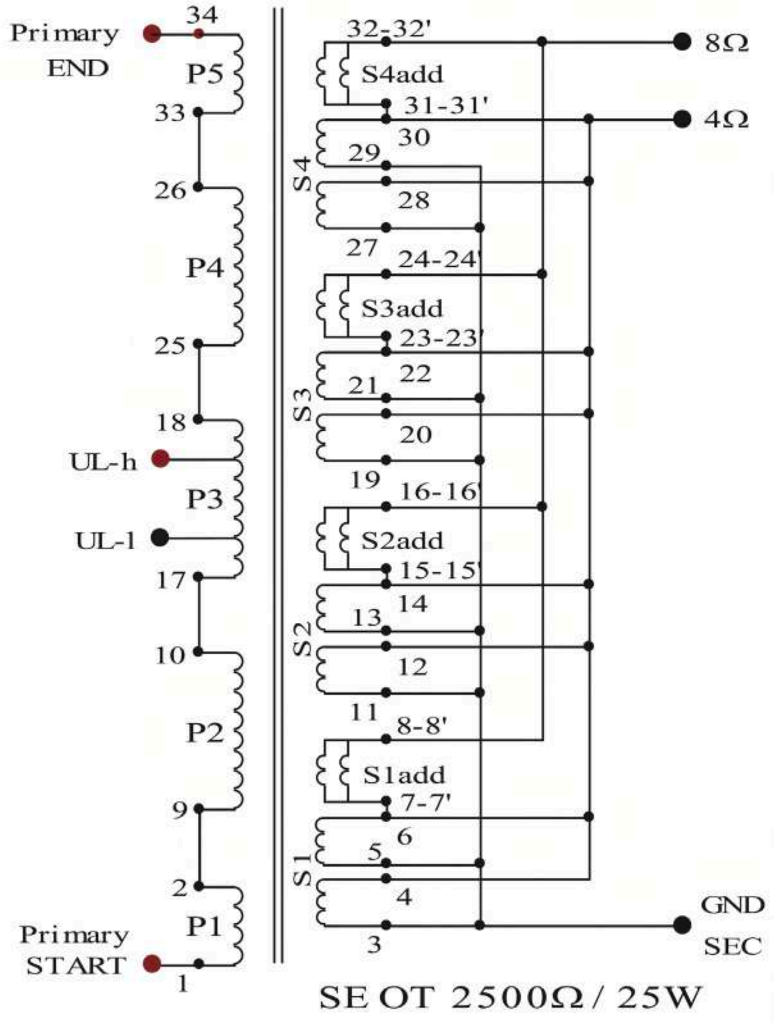
Air gap (production): δ = (0.4mm)

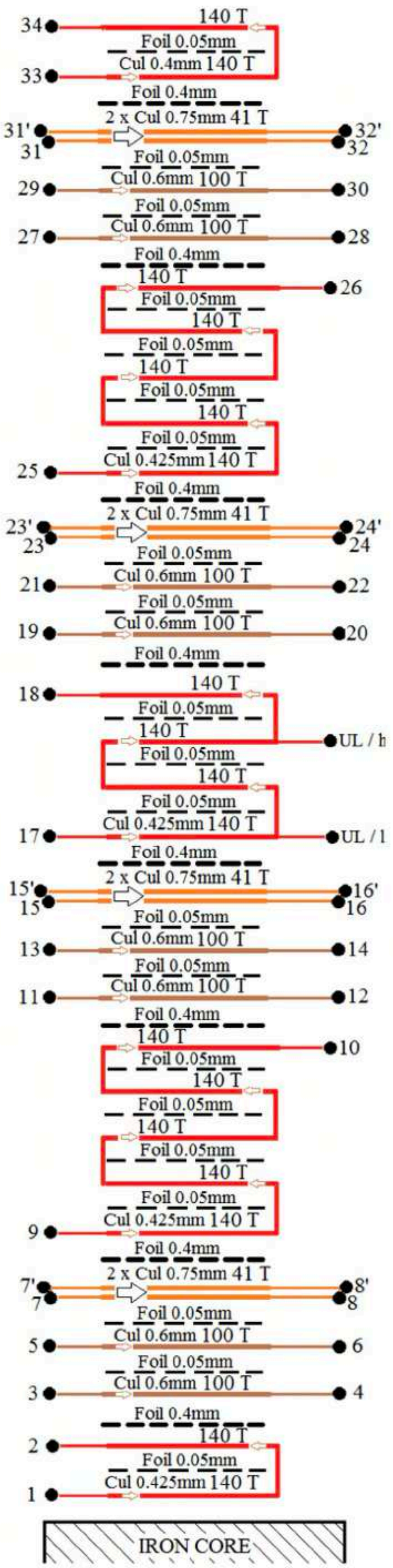
μ_e = 300
L_P = 20.7 H
L_S = 3.62 mH
$R_{P\text{-}DC}$ = 82.69 Ω
f_{C0} = 19.29 Hz
f_{sat} = 7.18 Hz
f_{-3dB} = 3.4 Hz
B_{DC} = 0.6137 T
B_{AC} = 0.4826 T

All **E** laminates stack in the same direction and all **I** laminates stack in a same direction.

 # Air gap between E and I laminate stacks: insulation foil # 0.4mm.

 # Ultra Linear Tap 44 %





PRIMARY

START	CONNECTED	END
1	2 - 9	34
	10 - 17	
	18 - 25	
	26 - 33	

UL / l UL / h

SECONDARY

COMMON

CONNECTED

3 - 5 - 11 - 13 - 19 - 21 - 27 - 29

4 Ω

CONNECTED

4 - 6 - (7-7') - 12 -14 - (15-15') - 20 - - 22 - (23-23') - 28 - 30 - (31-31')

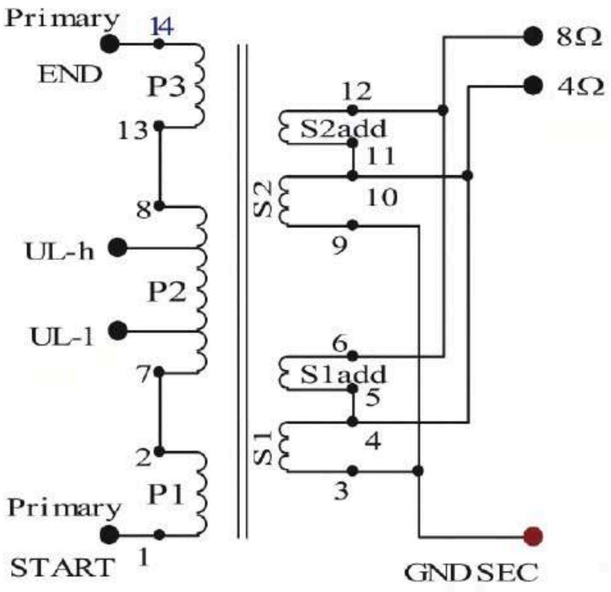
8 Ω

CONNECTED

(8-8') - (16-16') - (24-24') - (32-32')

SE OT 2500 / 4 / 8 / 25W

SE OUTPUT TRANSFORMER R_P = 5000 Ω / 15 W



SE OT 5000Ω / 15W

For example, it can be used in SE output stage with KT88 (Triode or Ultra linear)

Classic design

R_{La} = 5000 Ω; Load 4 Ω, 8 Ω; P_{max} = 15 W; I_{DC} = 90 mA

EI 120, 4% Si, Δ = 0.35 mm;
Coil Former: (40 x 50) mm

N_p = 3400 turns
N_{s4ohm} = 96 turns
N_{s8ohm} = 135 turns

Winding arrangement: 3P + 2S
(5 Primary sections – 4 Secondary sections)
(P1) – (S1, S1add) – (P2) – (S2, S2add) – (P3)

The total number of layers of the Primary winding: **20**

(5/20)P – ($S_{4\Omega}$+$S_{add\ 8\Omega}$) – (10/20)P – ($S_{4\Omega}$+$S_{add\ 8\Omega}$) – (5/20)P

Primary
Each section of the Primary consists of a certain number of layers of 170 turns of wire Cul Ø 0.265 mm (Ø 0.312 mm with insulation).
P1: 5 layers (850 turns), P2: 10 layers (1700 turns: 680 turns – UL TAP – 340 turns – UL TAP – 680 turns), P3: 5 layers (850 turns)
The Primary sections are connected in series (P1 + P2 + P3)

Secondary
4 Ω:
Each section of the 4 Ω consists of two layer of 48 turns (total: 96 turns) of wire Cul Ø 1mm (1.039 mm with insulation). Two sections of the 4Ω Secondary are connected in parallel ($S_{4\Omega}$ II $S_{4\Omega}$);
(S1 II S2): (96 turns II 96 turns)

8 Ω:
Each section of the additional Secondary winding consists of one layer of 39 turns of wires Cul Ø 1.25 mm (1.351 mm with insulation) Each section of the additional Secondary winding is connected in series with adjacent section of the 4Ω Secondary making 8 Ω Secondary sections. The sections of the 8Ω Secondary are connected in parallel.
($S_{4\Omega}$ + 39 turns) II ($S_{4\Omega}$+ 39 turns)

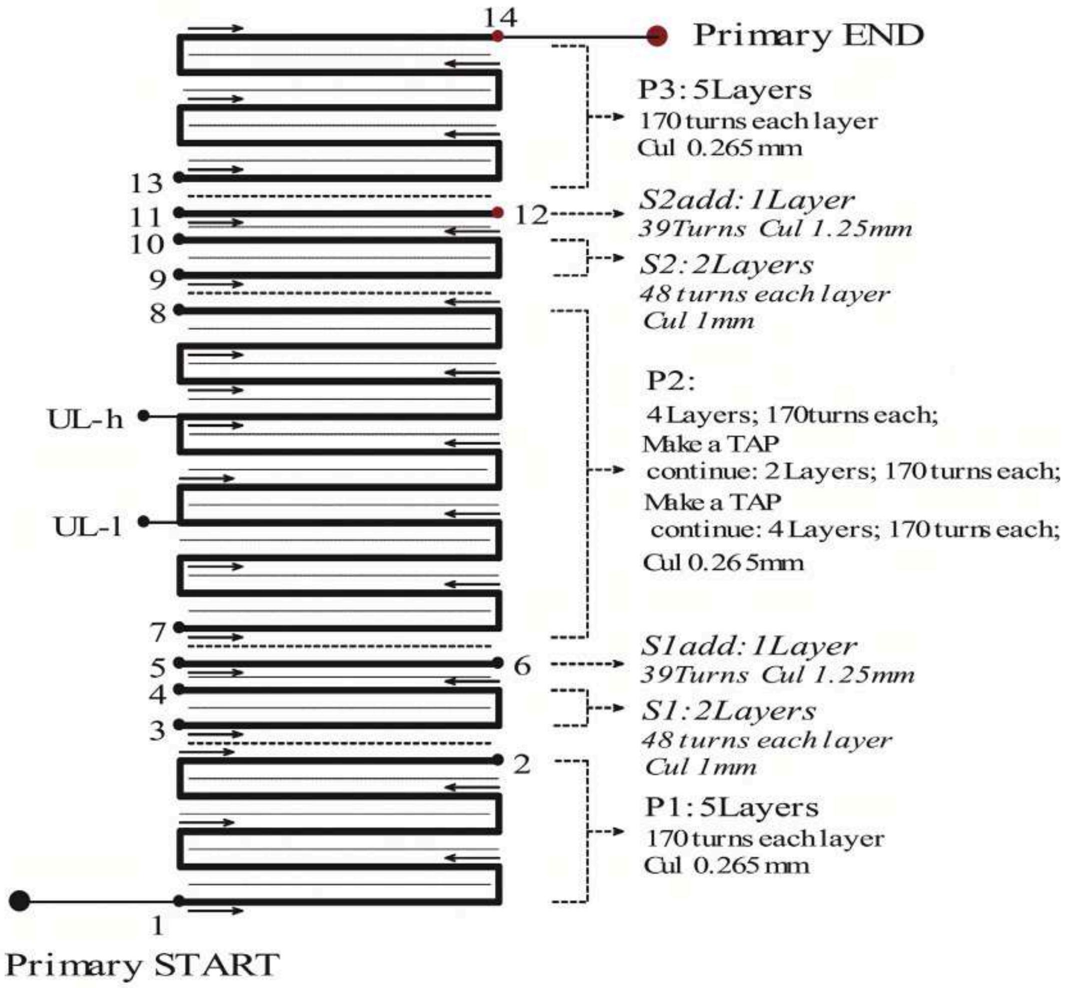
Insulation foil between P layers: 0.05 mm; Insulation foil between P and S layers: 0.4 mm;
Insulation foil between S layers: 0.05 mm

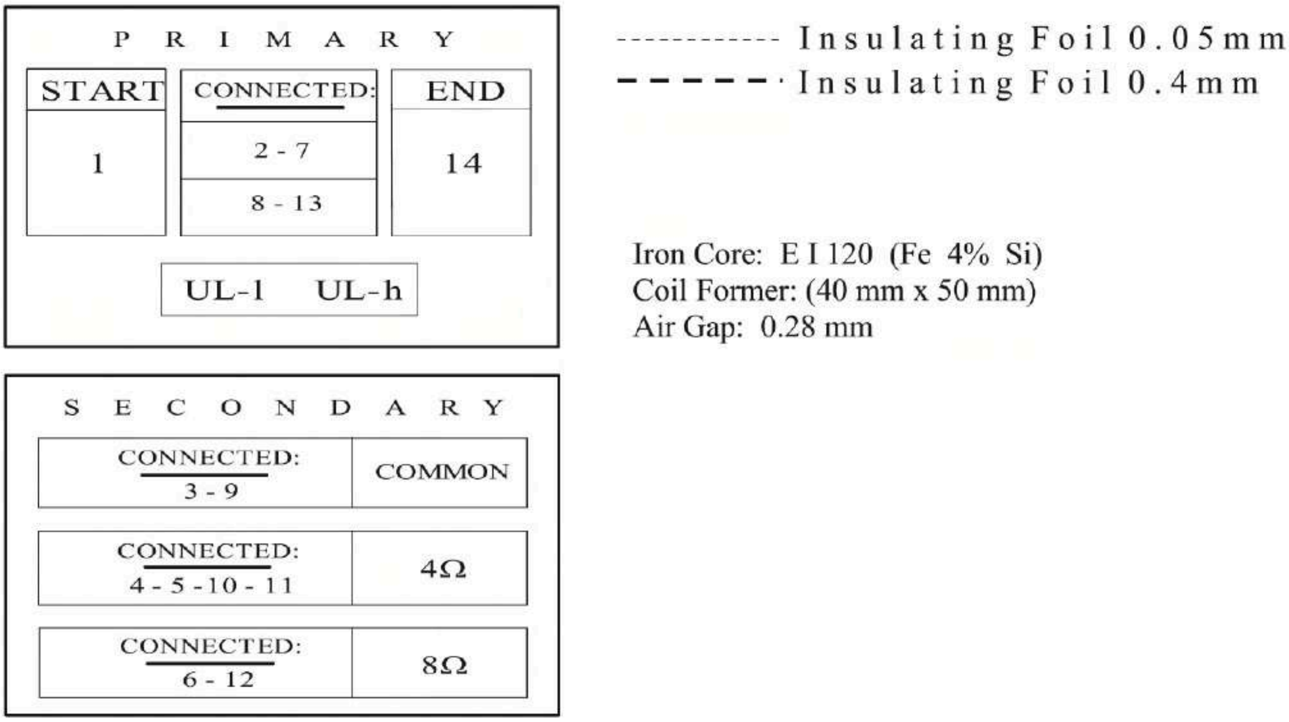
Air gap (production): δ = (0.28mm)

μ_e = 350
L_P = 43.9 H
L_S = 23.8 mH
$R_{P\text{-}DC}$ = 249 Ω
f_{C0} = 18 Hz
f_{sat} = 7.29 Hz
B_{DC} = 0.6054 T
B_{AC} = 0.5214 T
B_{DC} + B_{AC} = 1.12 T

All **E** laminates stack in the same direction and all **I** laminates stack in a same direction.
 # Air gap between E and I laminate stacks: insulation foil # 0.28mm.

 # Ultra Linear Tap 45 %





PRIMARY

START	CONNECTED:	END
1	2 - 7 8 - 13	14

UL-l UL-h

SECONDARY

CONNECTED:	
3 - 9	COMMON
4 - 5 -10 - 11	4Ω
6 - 12	8Ω

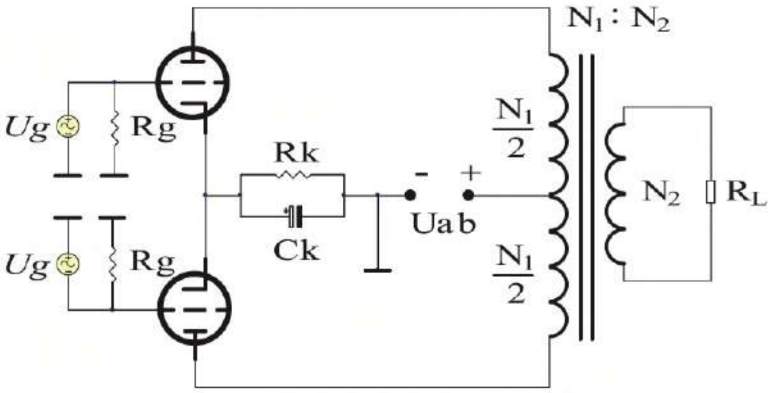
------------ Insulating Foil 0.05mm

– – – – – · Insulating Foil 0.4mm

Iron Core: E I 120 (Fe 4% Si)
Coil Former: (40 mm x 50 mm)
Air Gap: 0.28 mm

5.6 PUSH-PULL OUTPUT TRANSFORMER CALCULATIONS

Basic configuration of the Push-Pull output stage:



The magnetic fields caused by the DC quiescent currents of the output tubes flowing in opposite directions through the halves of the output transformer are cancelled out. Since there is no magnetic induction caused by DC current, the magnetic induction caused by AC current, which is also the total induction, can be higher than that of the SE transformer, but must be kept within the limits of the linear part of the B - H curve.

Maximum magnetic induction must be lower than 8000 Gauss (linear part of the B - H curve, low harmonic distortion).

E laminates are stacked one by one in the opposite direction. I laminates are inserted between E laminates

The procedure for calculating the Push Pull output transformer is similar to the procedure for calculating the Single Ended output transformer.

IRON CORE CROSS-SECTION

<u>First method:</u>

Calculation starts with some standard type of E/I transformer laminate, i.e. S_W .

$$S_{Fe}(cm^2) = \frac{1.8\times10^6\times P(W)}{f(Hz)\times B(Gauss)\times S_W(cm^2)\times J(\frac{A}{mm^2})} \qquad (501.48)$$

The above equation can be simplified using numerical values (**good performance output transformer)**:
f = (18 – 20) Hz, J = (1.5 – 2) A / mm^2, B = 8000 Gauss

$$S_{Fe}(cm^2) = \frac{(6.25 \div 8.3)\times P(W)}{S_W(cm^2)} \qquad (501.49)$$

B (Gauss) – Magnetic Flux Density
f (Hz) – The lowest frequency (Hz) at which the transformer can operate at full power
J (A/mm^2) – Electric Current Density
P (W) – Transformer power
S_W (cm^2) – Cross-section of the transformer core window (For E/I laminate $S_W = (a \times c)$ [cm^2]

If the proportion 1 ≤ h / b ≤ 1.6 is fulfilled, the choice of transformer laminate is good, but if the mentioned proportion is not fulfilled, the calculation process must be repeated with a new standard transformer laminate (S_W).

Second method:

$$S_{Fe(min)}(cm^2) = \sqrt{\frac{P \times G \times 10^6}{B \times J \times f}} \qquad (501.50)$$

P – Output power [W]
B – Induction **(8000 Gs)** (4% Si Ferro alloy transformer laminates)
G – Coefficient of weight ratio of iron core and cooper winding **(1.7 ... 2.5)**
J – Electric Current Density **(1.5 ... 2.5) A / mm^2**
f – Lowest frequency (Hz) at which the transformer can operate at full power

The above equation can be simplified using numerical values (**good performance output transformer)**:
f = (18 – 20) Hz, J = (1.5 – 2) A / mm^2, B = 8000 Gauss

$$S_{Fe(min)}(cm^2) = 2.95 \times \sqrt{P} \approx (3 - 3.4) \times \sqrt{P} \qquad (501.51)$$

IRON CORE (STANDARD) CHOICE

Choice of standard transformer laminate.

Choice of standard (or preferred) coil former.

Standard dimension of EI laminates tongue (b) and the height of the laminates stack (h)

$$S_{Fe(standard)} = b \times h \quad (cm^2)$$

The optimal choice is: b = h; (b / h) = 1

It is recommended to fulfill the ratio 1 ≤ h / b ≤ 1.6 .

It is recommended to choose b and h so that the condition $S_{Fe(st)} > S_{Fe}$ is fulfilled.

Calculation of the effective cross-section of the iron core .

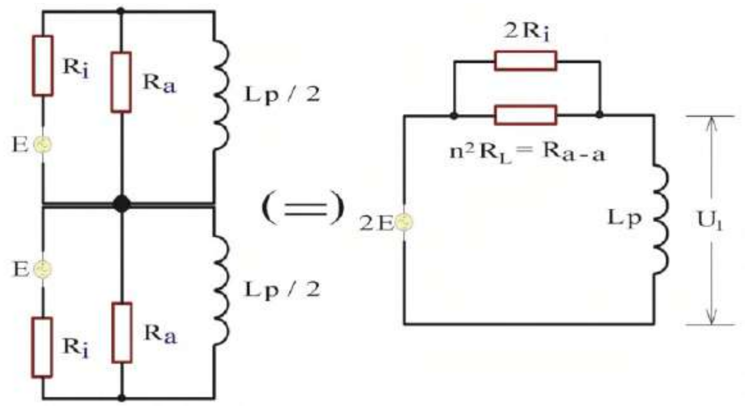
$$S_{Fe_{eff}} = (0.92 - 0.98) \times (b \times h) \qquad (cm^2)$$

Stacking coefficient of 4% Si non-oriented steel (NOSS) transformer laminate

1. thickness **0.35** mm **k = 0.96**
2. thickness **0.5** mm **k = 0.97**

The number of turns of the Primary:

Equivalent circuit (Class A, AB)



R_i – Internal resistance of the source (internal resistance of the tube)

R_a – Load Resistance of the Anode (Plate)
L_P – Inductance of the Primary
R_L – Load Resistance of the Secondary
b – Voltage attenuation (dB)

$f_{Low(-3dB)}$ – low cut-off frequency (Hz)

$$L = \frac{1}{2\times\pi\times f}\times\frac{R_i\times R_a}{R_i+R_a}\times\frac{1}{\sqrt{10^{0.1\times b}-1}},\ \text{and for b = 3dB} \rightarrow L = \frac{R_i\times R_a}{R_i+R_a}\times\frac{1}{2\times\pi\times f_{Low(-3dB)}}$$

A simplified equation for calculating the inductance of the Primary under the condition $R_{a-a} = 4 \times R_i$ and $f_{Low(-3dB)} = 2.5$Hz, or impedance of the Primary equal to the R_{a-a} at f = 15Hz:

$$L_P \geq \frac{R_{a-a}}{94.25}$$

The number of turns of the Primary:

$$L_P = \frac{1.256\times S_{eff}\times\mu_e\times N_P^2}{10^8\times l_{Fe}} \rightarrow N_P = \sqrt{\frac{2\times L[H]\times l_{Fe}[cm]\times 10^8}{1.256\times S_{eff}[cm^2]\times\mu_e}} \qquad (501.52)$$

$$N_P = 8923\times\sqrt{\frac{2\times L[H]\times l_{Fe}[cm]}{S_{eff}[cm^2]\times\mu_e}} \approx \mathbf{9000}\times\sqrt{\frac{2\times L[H]\times l_{Fe}[cm]}{S_{eff}[cm^2]\times\mu_e}} \qquad (501.53)$$

$\mu_e \approx$ **1000 (**as closely as possible stacked 4% Si Fe E and I laminates)

Class B Push Pull amplifier

The number of turns of the Primary

$$N_P \approx \mathbf{9000}\times\sqrt{\frac{4\times L[H]\times l_{Fe}[cm]}{S_{eff}[cm^2]\times\mu_e}} \qquad (501.54)$$

The number of turns of the Secondary:

$$N_S = N_P\times\sqrt{\frac{R_S}{R_{a-a}}} \qquad (501.55)$$

N_P – The number of turns of the Primary
N_S – The number of turns of the Secondary
R_{a-a} – Load resistance of the Primary (Ω)
R_S – Load Resistance of the Secondary (Ω)

Wire diameter of the Primary and Secondary windings:

Wire diameter of the Primary winding:

$$\boldsymbol{d_{Pmin}}(mm) = 1.13 \times \sqrt{\frac{I_a(A)}{J(\frac{A}{mm^2})}}$$

I_a –Plate (Anode) DC Current (**one tube**)

$$For\ J = 1.5\frac{A}{mm^2}: \quad \boldsymbol{d_{Pmin}}(mm) \approx 0.9 \times \sqrt{I_{P(DC)}(A)}$$

$$For\ J = 2\frac{A}{mm^2}: \quad \boldsymbol{d_{Pmin}}(mm) \approx 0.8 \times \sqrt{I_{P(DC)}(A)}$$

$$For\ J = 2.5\frac{A}{mm^2}: \quad \boldsymbol{d_{Pmin}}(mm) \approx 0.7 \times \sqrt{I_{P(DC)}(A)}$$

Selection of standard wire diameter: **$d_{P(standard)} \geq d_{Pmin}$**

The further calculation procedure is very similar to the single-ended output transformer calculation procedure explained above (adjusting N_P and N_S as an integer, winding arrangement, number of layers of the windings, turns per layer, checking the possibility of placing the windings in the window of the transformer core, choosing standard Cu wires, DC resistance of the Primary and Secondary windings, Cu losses, etc).

Saturation frequency of the iron core

$$\boldsymbol{f_{sat}} = \frac{22.6 \times U_{eff} \times 10^6}{S_{eff} \times N_p \times B_{max}}$$

B_{max} – maximal induction (For push-pull output transformer and 4% Si Ferro alloy: **Bmax = (16000 Gs**)

Checking L_P for actual number of turns of the Primary

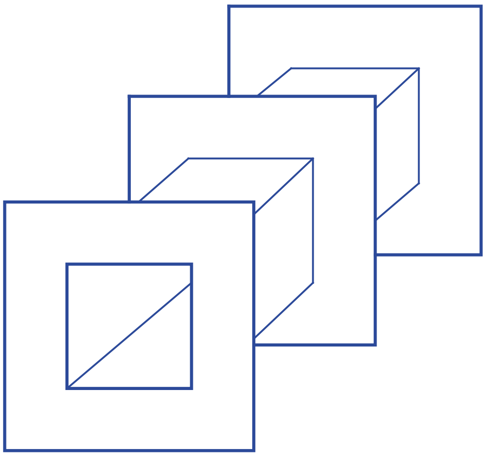
$$\boldsymbol{L_P} = \frac{1.256 \times S_{eff} \times \mu_e \times N_p^2}{10^8 \times l_{Fe}}$$

Calculating the low cut-off frequency

$$\boldsymbol{f_{Low(-3dB)}} = \frac{R_i \times R_a}{R_i + R_a} \times \frac{1}{2 \times \pi \times L_P}$$

The most important thing when designing a push-pull output transformer is to achieve electrical and magnetic symmetry of the transformer halves (symmetry of the primary windings halves). Minimizing the leakage inductance is also very important.

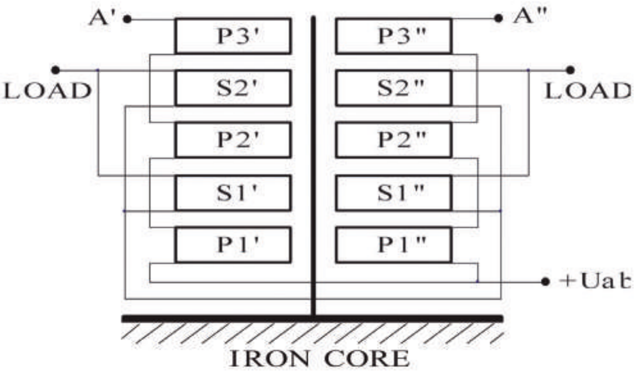
In practice, the technique of splitting the coil former into two equal parts is most often used:

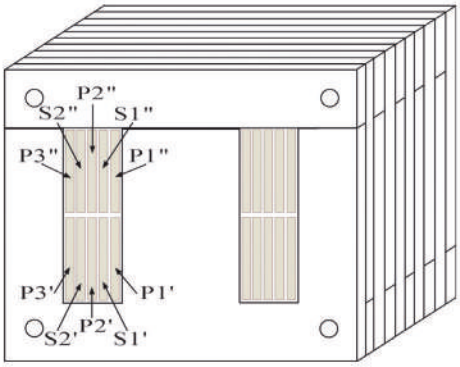


To make the transformer halves (primary and secondary windings) equal, usually one half of the transformer (half of the primary and secondary windings) is wound on one half of the coil former, then the coil former is rotated 180^0 and the other half of the transformer (half of the primary and secondary windings) is wound on the other half of the coil former in the same way as the first half is wound.

Example:

- Each half of the Primary winding is split into three equal sections connected in series. Between first and second Primary section, one section of the Secondary is inserted. Also, between second and third Primary section, one section of the Secondary is inserted. The Secondary sections are connected in parallel.
- The beginnings of the halves of the windings of the Primary (the beginnings of the first sections of the halves of the Primary) are connected together and this is the middle (CT - central or common) point of the Primary.
- The ends of the halves of the windings of the Primary (the ends of the third sections of the halves of the Primary) are the connection points for the anodes of the output tubes.
- Parallel-connected sections of the halves of the Secondary are connected in series.





Example1:

Push Pull Output transformer

$P_{OUT} = 20\ W$

$R_{(a\text{-}a)} = 5600\ \Omega$

 # For example, it can be used in PP output stage with 2 x WE300B ($U_{ak} = 350V$, $U_{gk} = 71V$, $I_a = 80mA$, $R_i = 720\ \Omega$ *)*

Cross-section of the transformer core

First method:

Calculation starts with EI 120 transformer laminate (cross-section of the core window: $S_W = (a \times c) = (2 \times 6) = 12\ cm^2$*)*

$$S_{Fe}(cm^2) = \frac{1.8\times 10^6 \times P(W)}{f(Hz)\times B(Gauss)\times S_W(cm^2)\times J\left(\frac{A}{mm^2}\right)} = \frac{1.8\times 10^6\times 20W}{18Hz\times 8000Gs\times 12(cm^2)\times 1.5\left(\frac{A}{mm^2}\right)} = 13.89\ cm^2$$

$$h = \frac{S_{Fe}}{b} = \frac{13.89}{4} = 3.47cm$$

Standard coil former EI 120: (b x h) = ***(4 cm x 4 cm)***

Effective cross-section of the transformer core *(EI 120 Fe 4% Si laminate Δ = 0.35mm, stacking coefficient k = 0.96)*

$$\boldsymbol{S_{Fe\ eff}} = k\times(b\times h) = 0.96\times(4\times 4) = \mathbf{15.36\ cm^2}$$

Second method:
B (Gauss) – Magnetic Flux Density
f (Hz) – The lowest frequency (Hz) at which the transformer can operate at full power
J (A/mm^2) – Electric Current Density
P (W) – Transformer power
G **–** *Coefficient of weight ratio of iron core and cooper winding (1.7 ÷ 2.5)*

$$S_{Fe(min)}(cm^2) = \sqrt{\frac{P\times G\times 10^6}{B\times J\times f}} = \sqrt{\frac{20\times 2.5\times 10^6}{8000\times 1.5\times 18}} = 15.21\ cm^2$$

*Height of stack of transformer laminates (**h**):*

$$h = \frac{S_{Fe}}{b} = \frac{15.21}{4} = 3.8\ cm$$

Standard coil former EI 120: (b x h) = ***(4 cm x 4 cm)***

Effective cross-section of the transformer core *(EI 120 Fe 4 % Si laminate Δ = 0.35 mm, stacking coefficient k = 0.96)*

$$\boldsymbol{S_{Fe\ eff}} = k\times(b\times h) = 0.96\times(4\times 4) = \mathbf{15.36\ cm^2}$$

Conclusion:
Coil Former: (4 cm x 4 cm)
EI 120 4 % Si Fe transformer laminate, Δ = 0.35 mm
$S_{Fe\ eff}$ = 15.36 cm^2

The number of turns of the Primary:

*Inductance of the (1 / 2) Primary, $\mathbf{L_{P/2}}$ **:***

$$R_a = \frac{R_{a-a}}{2} = \frac{5600\ \Omega}{2} = 2800\ \Omega$$

b – Voltage attenuation (dB) (for example: b = - 0.1dB at f = 20Hz)

$$\boldsymbol{L_{P/2}} = \frac{1}{2\times\pi\times f}\times\frac{R_i\times R_a}{R_i+R_a}\times\frac{1}{\sqrt{10^{0.1\times b}-1}} = \frac{1}{2\times\pi\times 20}\times\frac{720\ \Omega\times 2800\ \Omega}{720\ \Omega+2800\ \Omega}\times\frac{1}{\sqrt{10^{0.1\times 0.1}-1}} = \mathbf{29.88\ H\approx 30\ H}$$

Total inductance of the Primary: $\mathbf{L_P}$ = 2 x $L_{P/2}$ = 2 x 29.88 H ≈ ***60 H***

Quick calculation of the total inductance of the Primary: $\boldsymbol{L_P} \geq \frac{R_{a-a}}{94.25} \geq \frac{5600}{94.25} \geq \mathbf{59\ H}$

Total number of turns of the Primary:

$$N_P = 8923 \times \sqrt{\frac{2 \times L_{P/2}[H] \times l_{Fe}[cm]}{S_{eff}[cm^2] \times \mu_e}} = 8923 \times \sqrt{\frac{2 \times 29.88 \times 5.57 \times 4}{15.36 \times 1000}} = 2627.1\ Turns$$

Wire diameter of the Primary winding (for $J = 1.5A/mm^2$):

$$d_{Pmin}(mm) = 1.13 \times \sqrt{\frac{I_a(A)}{J\left(\frac{A}{mm^2}\right)}} = 1.13 \times \sqrt{\frac{0.08}{1.5}} = 0.26\ mm$$

Maximum wire diameter of the Primary winding (for N_P =2628 turns):

$$d_{Wp}(mm) = \sqrt{\frac{0.28 \times S_W(mm^2)}{N_P}} = \sqrt{\frac{0.28 \times 1200}{2628}} = 0.357\ mm$$

Standard wire diameter(d_P): ***CuI Ø 0.3mm (Ø 0.355mm with insulation).***

Number of turns of the Secondary (R_L = 4 Ω and R_L = 8 Ω):

$$N_{S(4\Omega)} = N_P \times \sqrt{\frac{R_L}{R_{a-a}}} = 2628 \times \sqrt{\frac{4}{5600}} = \mathbf{70.2\ Turns}$$

Number of turns (INTEGER) of the Secondary (R_L = 4 Ω) = 70 Turns

Number of turns of the Secondary (R_L = 8 Ω):

$$N_{S(8\Omega)} = N_P \times \sqrt{\frac{R_L}{R_{a-a}}} = 2628 \times \sqrt{\frac{8}{5600}} = \mathbf{99.3\ Turns}$$

Number of turns (INTEGER) of the Secondary (R_L = 8Ω) = 99 Turns

Saturation frequency of the Ferromagnetic core

$$f_{sat} = \frac{22.6 \times U_{eff} \times 10^6}{S_{eff} \times N_p \times B_{max}} = \frac{22.6 \times \sqrt{P_{OUT}[W] \times R_{a-a}[\Omega]} \times 10^6}{S_{eff} \times N_p \times B_{max}} = \frac{22.6 \times \sqrt{20 \times 5600} \times 10^6}{15.36 \times 2628 \times 16000} = \mathbf{11.7\ Hz}$$

$\mathbf{B_{max}}$ *– maximal induction (For push-pull output transformer and 4% Si Ferro alloy:* $\mathbf{B_{max}}$ ***= (16000 Gs)****)*

Checking L_P for actual number of turns of the Primary

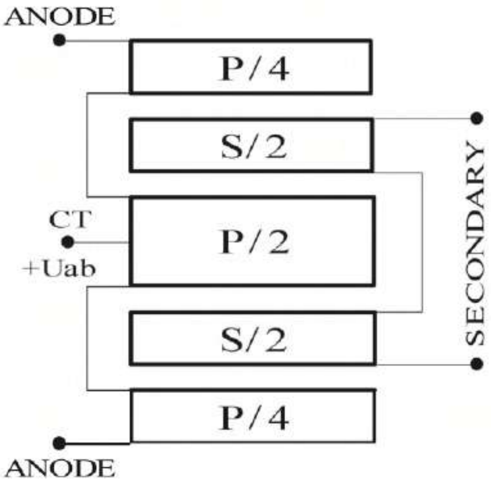
$$L_P = \frac{1.256 \times S_{eff} \times \mu_e \times N_p^2}{10^8 \times l_{Fe}} = \frac{1.256 \times 15.36 \times 1000 \times 2628^2}{10^8 \times 5.57 \times 4} = 59.8\ H$$

Calculating the low cut-off frequency

$$f_{Low(-3dB)} = \frac{R_i \times R_a}{R_i + R_a} \times \frac{1}{2 \times \pi \times L_P} = \frac{720 \times 2800}{720 + 2800} \times \frac{1}{2 \times \pi \times (\frac{59.8}{2})} = \mathbf{3\ Hz}$$

Winding arrangement

First design: a textbook example of transformer design
The simplest, but very effective winding arrangement is:



First, it is necessary to adjust the number of turns of the Primary and Secondary (integer).
The calculated diameter of the Cu wire of the Primary is 0.3 mm, (standard: insulated Cul Ø 0.355 mm).

The maximum number of turns per layer of the Primary winding is:

$$Turn/_{Layer} = \frac{0.97 \times c''}{d_{Wp(standard)}} = \frac{0.97 \times (60-4)}{d_{Wp(standard)}\,(mm)}$$

$$= \frac{54.32}{0.355\,(mm)} = 153$$

Number of layers:

N_P / (Turn/layer) = 2628 / 153 = 17.17

Number of layers (First higher INTEGER) = ***18***

New maximum number of turns of the Primary is: (Turn / layer) x Number of layers = 153 x 18 = 2754

Since the actual number of turns of the Primary must be less than the calculated maximum number of turns of the Primary:

Number of turns of the Primary, N_P = 2700 Turns

Number of turns of the 4 Ω Secondary, $N_{S\,4\Omega}$:

$$N_{S(4\Omega)} = N_P \times \sqrt{\frac{R_L}{R_{a-a}}} = 2700 \times \sqrt{\frac{4}{5600}} = 72.16\,Turns$$

Number of turns of the 4 Ω Secondary, $N_{S\,4\Omega}$ INTEGER: 72 Turns

Number of turns of the 8 Ω Secondary, $N_{S\,8\Omega}$ INTEGER: 101 Turns

Winding arrangement:
18 layers of the Primary winding can be arranged as follows:

(1/4)Primary – (1/2) Secondary – (1/4)Primary – TAP (CT) – (1/4) Primary – (1/2) Secondary – (1/4)Primary

*or: (**4/18)P – S / 2 – (5/18)P – TAP (CT) – (5/18)P – S / 2 – (4/18)P***

Winding arrangement of the Secondary:

Number of turns of one half of the 4 Ω Secondary: ***$N_{S(4\Omega)}$ / 2 = 72 / 2 = 36 Turns***
It can be wound in one layer with the Cu wire (insulated) of the maximum diameter:

$$d_{WS(4\Omega)}(mm) = \frac{0.97 \times c''}{Turn/_{Layer}} = \frac{0.97 \times (60-4)}{Turn/_{Layer}} = \frac{54.32}{36} = \mathbf{1.508\,mm}$$

Standard wire : ***Cul Ø 1.4mm (insulated Cul Ø 1.506 mm)***

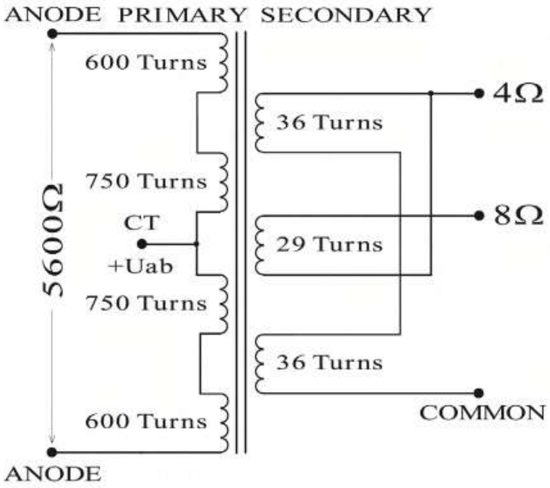
8Ω Secondary: additional winding connected in series with 4Ω Secondary winding. As an example, two winding method are explained:

1. *Number of turns of the additional winding is:* $N_{S\,8\Omega} - N_{S\,4\Omega}$ *= 101 turns - 72 turns = 29 Turns*

   It can be wound in one layer with the Cu wire (insulated) of the maximum diameter:

$$d_{WS(8\Omega)}(mm) = \frac{0.97 \times c^{"}}{Turn/_{Layer}} = \frac{0.97 \times (60-4)}{Turn/_{Layer}} = \frac{54.32}{29} = \mathbf{1.873}\ \boldsymbol{mm}$$

Standard wire: ***CuI Ø 1.7mm (insulated CuI*** Ø 1.813mm)



2. Half of the 8 Ω Secondary is:

   1 / 2 ($N_{S\,8\Omega}$ INTEGER =101 Turns) = 50.5 Turns

   An additional winding connected in series with half of the 4 Ω Secondary winding *is:*

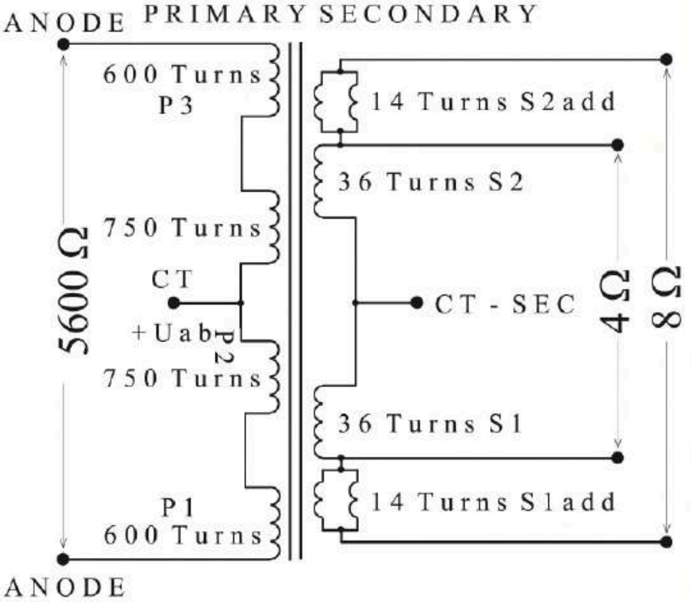
   50.5 turns - (72 / 2) turns = 14.5 Turns.

   Half of the additional winding (INTEGER): 14 Turns

   It can be wound in one layer with the Cu wire (insulated) of the maximum diameter:

$$d_{Wadd} = \frac{0.97 \times c^{"}}{\left(Turn/_{Layer}\right)} = \frac{0.97 \times (60-4)}{14} = 3.88\ mm$$

3.88mm diameter Cu wire is very thick and difficult to wind. To avoid this problem, an additional winding (14 Turns) can be wound with two CuI Ø 1.8 mm (insulated CuI Ø 1.916 mm) wires (bifilar: parallel-winding, parallel connection).



In any case, it is necessary to check the total radial dimension of the windings including the insulation between the layers of the Primary winding, between the layers of the Secondary winding and between the layers of the Primary and Secondary windings.

Insulation between the layers of the Primary windings:
Δ_i *= 0.05 mm*

Insulation between the layers of the Secondary winding:
Δ_i *= 0.05 mm*

Insulation between the layers of the Primary and Secondary windings:

Δ_{ii} *= (0.3 ÷ 0.4) mm*

$$\boldsymbol{height_{total}} = n_{Players} \times (d_{P\,with\,isolation} + \Delta_i) + n_{Slayers} \times (d_{S\,with\,isolation} + \Delta_i) + n_{P-S\,layers} \times \Delta_{ii}$$

1. $\boldsymbol{height_{total}} = \boldsymbol{h_0} = 18 \times (0.355 + 0.05) + 2 \times (1.506 + 0.05) + 1 \times (1.813 + 0.05) + 6 \times 0.4 = \mathbf{14.665}\ mm$

2. $\boldsymbol{height_{total}} = \boldsymbol{h_0} = 18 \times (0.355 + 0.05) + 2 \times (1.506 + 0.05) + 2 \times (1.916 + 0.05) + 4 \times 0.4 = \mathbf{15.934}\ mm$

 # Condition: $h_0 < h_c = a - s$ must be fulfilled: 14.665 ≤ 20 – 2.5 = 17.5 mm and 15.934 ≤ 20 – 2.5 = 17.5 mm

*Calculation of **mean length per turn (ℓ_t)**:*

$$\ell_t = 2 \times h + \pi \times h_0 + 2 \times b + 4 \times s = 2 \times 40 + 3.14 \times 15.934 + 2 \times 40 + 4 \times 2 = 218\ mm = 0.218\ m$$

Wire length of the primary winding:

$$\ell_{wP} = N_P \times \ell_t = 2700 \times 0.218 = 588.6\ m$$

Wire length of the 4 Ω secondary:

$$\ell_{w4_\Omega} = N_{S4_\Omega} \times \ell_t = 72 \times 0.218 = 15.69\ m$$

Wire length of the additional winding:

$$\ell_{wadd} = N_{S8add_\Omega} \times \ell_t = 29 \times 0.218 = 6.3\ m$$

DC resistance of the winding ($R_{c\text{-}dc}$):

$$R_{c-DC} = \rho \times \frac{4 \times \ell_w}{\pi \times d^2}$$

ρ – *specific resistance of the wire* ***(Cu wire,*** $\boldsymbol{\rho = 0.0175\ \Omega \frac{mm^2}{m}}$ ***)***

DC resistance of the Primary winding: $\boldsymbol{R_{P-DC}} = \rho \times \frac{4 \times \ell_w}{\pi \times d^2} = 0.0175 \times \frac{4 \times 588.6}{3.14 \times 0.3^2} = \mathbf{145.7\ \Omega}$

DC resistance of the 4Ω Secondary winding: $\boldsymbol{R_{4\Omega-DC}} = \rho \times \frac{4 \times \ell_w}{\pi \times d^2} = 0.0175 \times \frac{4 \times 15.69}{3.14 \times 1.4^2} = 0.177\ \Omega$

DC resistance of the additional winding:

1. $\boldsymbol{R_{add-DC}} = \left(\rho \times \frac{4 \times \ell_w}{\pi \times d^2}\right) = (0.0175 \times \frac{4 \times 6.3}{3.14 \times 1.7^2}) = 0.0486\ \Omega$

   An additional winding is connected in series with the 4Ω Secondary section, so the DC resistance of one 8Ω Secondary section is: $R_{8\Omega\text{ - }DC} = R_{4\Omega\text{ - }DC} + R_{add} = 0.177\ \Omega + 0.0486\ \Omega = 0.225\ \Omega$.

2. $\boldsymbol{R_{add-DC}} = \frac{1}{2} \times \left(\rho \times \frac{4 \times \ell_w}{\pi \times d^2}\right) = \frac{1}{2} \times (0.0175 \times \frac{4 \times 6.3}{3.14 \times 1.8^2}) = 0.021\ \Omega$

   An additional winding is connected in series with the 4Ω Secondary section, so the DC resistance of one 8Ω Secondary section is: $R_{8\Omega\text{ - }DC} = R_{4\Omega=dc} + R_{8\Omega\ add} = 0.177\ \Omega + 0.021\ \Omega = 0.198\ \Omega$.

Calculation and checking of copper (Cu) losses in transformer windings:

***Cu loss in Primary winding*:**

$$\boldsymbol{P_{P-Cu\ loss}}(\%) = 100 \times \frac{R_{P(c-DC)}}{R_{P(c-DC)} + R_{La}} = 100 \times \frac{145.7}{145.7 + 5600} = \mathbf{2.53}\ \%$$

Cu loss in Secondary winding:

$$\boldsymbol{P_{S4\Omega-Cu\ loss}}(\%) = 100 \times \frac{R_{S(c-DC)}}{R_{S(c-DC)} + R_S} = 100 \times \frac{0.177}{0.177 + 4} = \mathbf{4.2}\ \%$$

Cu loss in 8Ω Secondary winding:

$$\boldsymbol{P_{S8\Omega-Cu\,loss}}(\%) = 100 \times \frac{R_{S(c-DC)}}{R_{S(c-DC)} + R_S} = 100 \times \frac{0.225}{0.225 + 8} = \mathbf{2.7}\ \%$$

and

$$\boldsymbol{P_{S8\Omega-Cu\,loss}}(\%) = 100 \times \frac{R_{S(c-DC)}}{R_{S(c-DC)} + R_S} = 100 \times \frac{0.198}{0.198 + 8} = \mathbf{2.4}\ \%$$

Total Cu loss (for 4Ω Secondary):

$$\boldsymbol{P_{loss}}(\%) = \boldsymbol{P_{P-Cu\,loss}}(\%) + \boldsymbol{P_{S-loss}}(\%) = \mathbf{2.53 + 4.2 = 6.73}\ \% < 10\ \%$$

Total Cu loss (for 8Ω Secondary):

$$\boldsymbol{P_{cu\,loss}}(\%) = \boldsymbol{P_{P-cu\,loss}}(\%) + \boldsymbol{P_{S-cu\,loss}}(\%) = \mathbf{2.53 + 2.7 = 5.23}\ \% < 10\ \%$$

and

$$\boldsymbol{P_{Cu\,loss}}(\%) = \boldsymbol{P_{P-Cu\,loss}}(\%) + \boldsymbol{P_{S-Cu\,loss}}(\%) = \mathbf{2.53 + 2.4 = 4.93}\ \% < 10\ \%$$

Check the inductance of the primary winding (L_P):

$$\boldsymbol{L_P} = \frac{1.256 \times S_{eff} \times \mu_e \times N_p^2}{10^8 \times l_{Fe}} = \frac{1.256 \times 15.36 \times 1000 \times 2700^2}{10^8 \times 5.57 \times 4} = \mathbf{63\ H}$$

Calculating $f_{low\,(-3dB)}$ (for calculated L_P):

$$\boldsymbol{R_A} = \frac{(2 \times R_i + R_{P-DC}) \times R_{La-a}}{2 \times R_i + R_{P-DC} + R_{La-a}} = \frac{(2 \times 720 + 145.7) \times 5600}{2 \times 720 + 145.7 + 5600} = \mathbf{1235.7\ \Omega}$$

$$\boldsymbol{f_{low}} = \frac{R_A}{2 \times \pi \times L_P} = \frac{1235.7}{2 \times 3.14 \times 63} = \mathbf{3.12\ Hz}$$

Calculating f_{Co} :

$$\boldsymbol{f_{C_0}} = \frac{R_{La-a}}{2 \times \pi \times L_p} = \frac{5600}{2 \times 3.14 \times\times 63} = \mathbf{14.15\ Hz}$$

Calculating saturation frequency f_{sat} :

$$\boldsymbol{f_{sat}} = \frac{22.6 \times U_{eff} \times 10^6}{S_{eff} \times N_p \times B_{max}} = \frac{22.6 \times \sqrt{P_{OUT}[W] \times R_{a-a}[\Omega]} \times 10^6}{S_{eff} \times N_p \times B_{max}} = \frac{22.6 \times \sqrt{20 \times 5600} \times 10^6}{15.36 \times 2700 \times 16000} = \mathbf{11.4\ Hz}$$

***B_{max}** – maximal induction (For push-pull output transformer and 4 % Si Ferro alloy: **B_{max} = (16000 Gs)**)*

Calculation of the lowest frequency at which the induction does not exceed 8000 Gauss (linear part of the BH curve – low harmonic distortion):

$$f_{(8000\,Gs)} = \frac{U_{eff} \times 10^8}{4.44 \times \boldsymbol{B_{AC}} \times N_p \times S_{Fe_{eff}}} = \frac{10^8 \times \sqrt{P_{OUT} \times R_{La}}}{4.44 \times \boldsymbol{B_{AC}} \times N_p \times S_{Fe_{eff}}} = \frac{10^8 \times \sqrt{20 \times 5600}}{4.44 \times 8000 \times 2700 \times 15.36} = \mathbf{22.7\ Hz}$$

B_{AC} (at 20Hz) ≈ 9000 Gauss

Leakage inductance and high cut-off frequency:

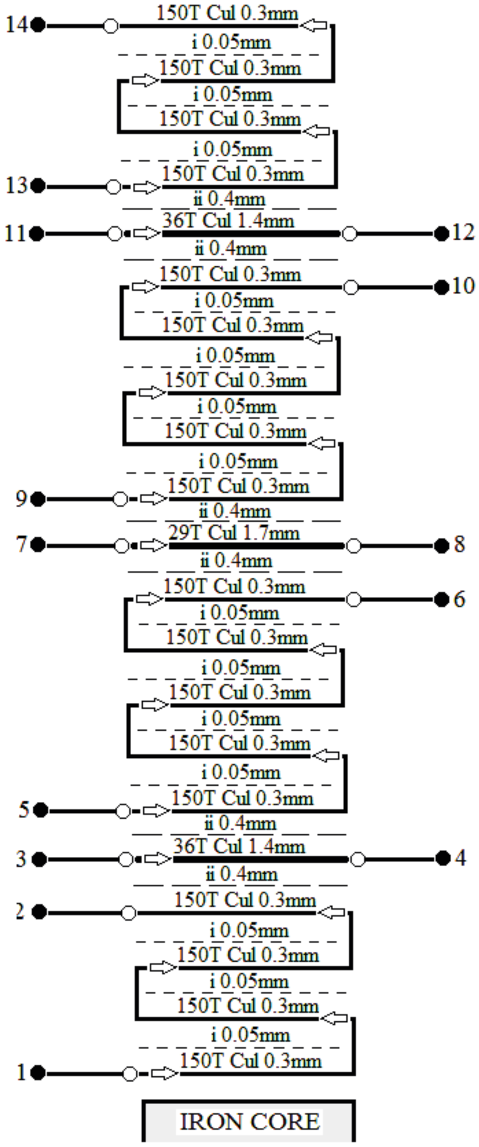
$$L_S = \frac{0.417 \times Np^2 \times l_t \times [(2 \times n_{P-S} \times \Delta_{ii}) + h_{total}]}{10^9 \times n_{P-D}^2 \times c"}$$

$$L_S = \frac{0.417 \times 2700^2 \times 178 \times [(2 \times 4 \times 0.4) + 15.934]}{10^9 \times 4^2 \times 56} = \mathbf{0.011H = 11\ mH}$$

$$R_B = 2 \times R_i + R_{P-DC} + R_{La-a} = 2 \times 720\Omega + 145.7\Omega + 5600\Omega = \mathbf{7185.7\ \Omega}$$

$$f_{(-3dB)high} = \frac{R_B}{2\pi \times L_S} = \frac{7185.7}{2 \times 3.14 \times 0.011} = \mathbf{104019\ Hz}$$

In this example, a simple ("classic") construction of a Push Pull output transformer is described. A standard EI 120 coil former (4 x 4) cm is used which can be easily found on the market. Other materials required for making transformer (transformer laminates, Cu wires and insulating foils) are standard and also easily available on the market. The winding process is also simple. Despite the simplicity of the construction, the performance of this PP output transformer is very good.



Iron Core: E / I 120
Laminate: 0.35mm Fe 4% Si
Coil Former: (40 x 40) mm

P1: (1 - 2), 4 Layers, 150 Turns per layer
(Total: 600 Turns), Cul 0.3mm
Insulating Foil 0.05mm between layers

i i : Insulating Foil 0.4mm

S1: (3 - 4). 1 Layer 36 Turns Cul 1.4mm

i i : Insulating Foil 0.4mm

P2 / 2: (5 - 6), 5 Layers, 150 Turns per layer
(Total: 750 Turns), Cul 0.3mm
Insulating Foil 0.05mm between layers

i i : Insulating Foil 0.4mm

Sadd: (7 - 8), 1 Layer 29 Turns Cul 1.7mm

i i : Insulating Foil 0.4mm

P2 / 2: (9 - 10), 5 Layers, 150 Turns per layer
(Total: 750 Turns), Cul 0.3mm
Insulating Foil 0.05mm between layers

i i : Insulating Foil 0.4mm

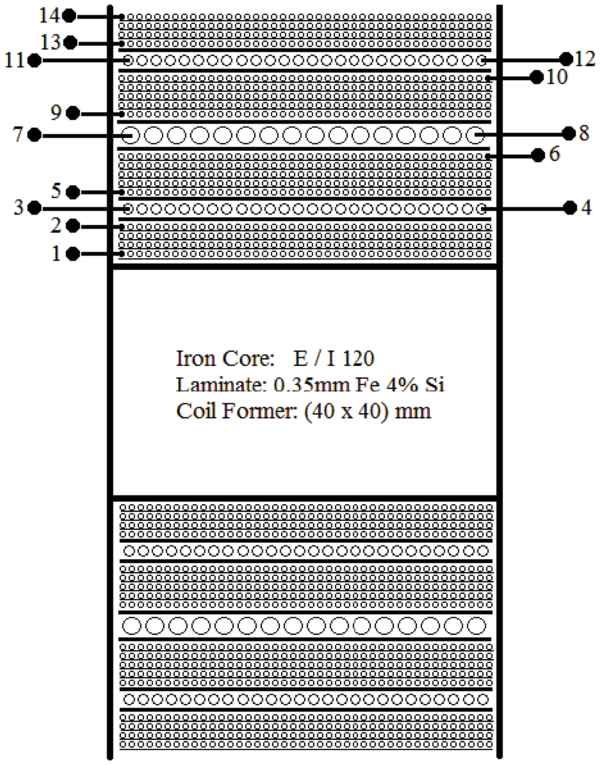
S2: (11 - 12), 1 Layer 36 Turns Cul 1.4mm

i i : Insulating Foil 0.4mm

P3: (13 - 14). 4 Layers, 150 Turns per layer
(Total: 600 Turns), Cul 0.3mm
Insulating Foil 0.05mm between layers

	PRIMARY
A	1
	CONNECT (2 - 5)
Uab	CONNECT (6 - 9)
	CONNECT (10 - 13)
A	14

	SECONDARY
COMMON	3
	CONNECT (4 - 11)
4 Ω	CONNECT (7 - 12)
8 Ω	8



P1: (1 - 2), 4 Layers, 150 Turns per layer
(Total: 600 Turns), Cul 0.3mm
Insulating Foil 0.05mm between layers

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Insulating Foil 0.05mm between layers

i i : Insulating Foil 0.4mm

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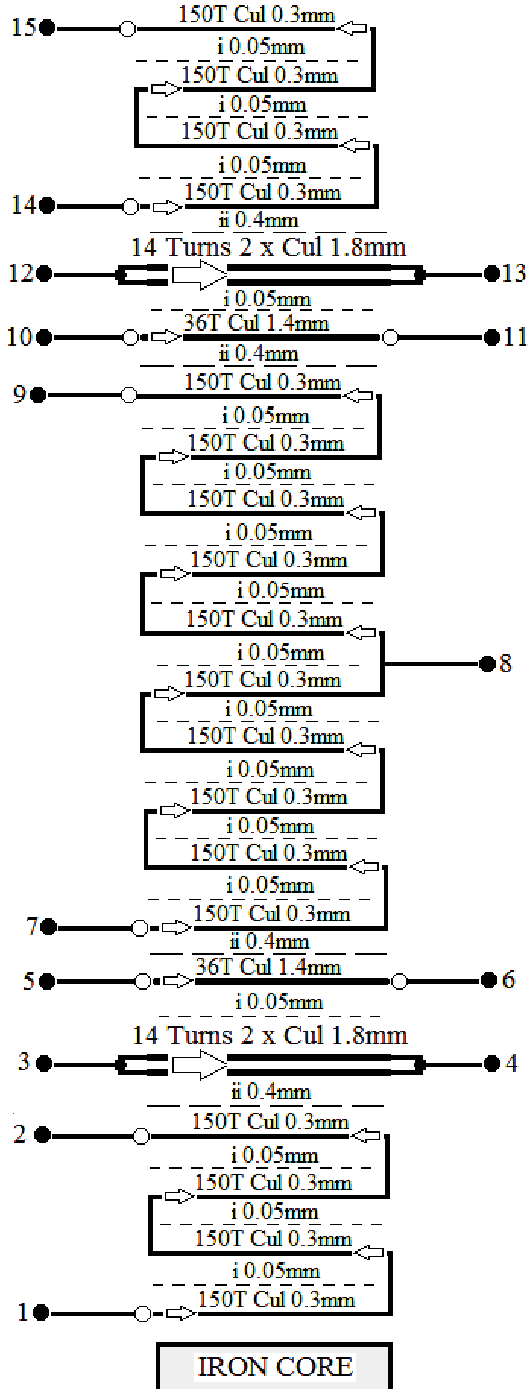
(Total: 600 Turns), Cul 0.3mm
Insulating Foil 0.05mm between layers

	PRIMARY
A	1
	CONNECT (2 - 5)
Uab	CONNECT (6 - 9)
	CONNECT (10 - 13)
A	14

	SECONDARY
COMMON	3
	CONNECT (4 - 11)
4 Ω	CONNECT (7 - 12)
8 Ω	8



Iron Core: E / I 120
Laminate: 0.35mm Fe 4% Si
Coil Former: (40 x 40) mm

P1: (1 - 2), 4 Layers, 150 Turns per layer
(Total: 600 Turns), Cul 0.3mm
Insulating Foil 0.05mm between layers

i i : Insulating Foil 0.4mm

S1add: (3 - 4), 1 Layer 14 Turns 2 x Cul 1.8mm

i: Insulating Foil 0.05mm

S1: (5 - 6), 1 Layer 36 Turns Cul 1.4mm

i i : Insulating Foil 0.4mm

P2: (7 - 8 - 9), 10 Layers (5 + 5), 150 Turns per layer
(Total: 1500 (750 + 750) Turns), Cul 0.3mm
(5 Layers, TAP(8), 5 Layers)
Insulating Foil 0.05mm between layers

i i : Insulating Foil 0.4mm

S2: (10 - 11), 1 Layer 36 Turns Cul 1.4mm

i: Insulating Foil 0.05mm

S2add: (12 - 13), 1 Layer 14 Turns 2 x Cul 1.8mm

i i : Insulating Foil 0.4mm

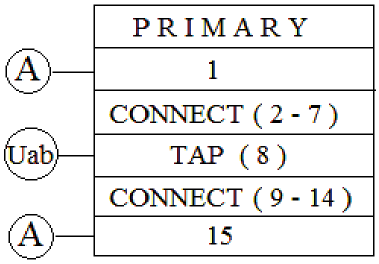
P2 / 2: (9 - 10), 5 Layers, 150 Turns per layer
(Total: 750 Turns), Cul 0.3mm
Insulating Foil 0.05mm between layers

i i : Insulating Foil 0.4mm

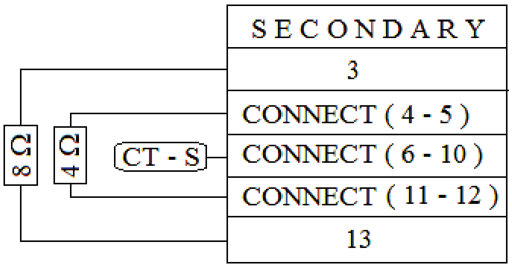
S2: (11 - 12), 1 Layer 36 Turns Cul 1.4mm

i i : Insulating Foil 0.4mm

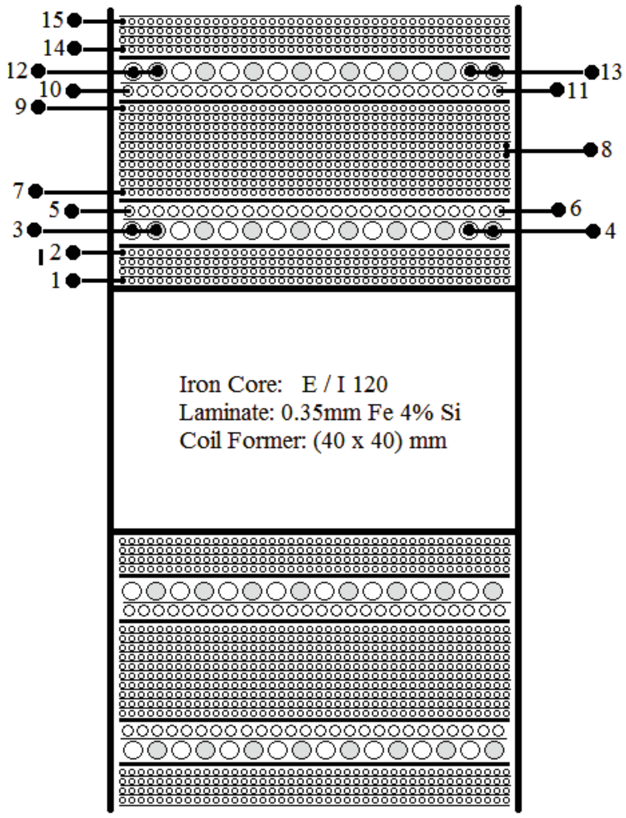
P3: (14 - 15), 4 Layers, 150 Turns per layer
(Total: 600 Turns), Cul 0.3mm
Insulating Foil 0.05mm between layers



PRIMARY
1
CONNECT (2 - 7)
TAP (8)
CONNECT (9 - 14)
15



SECONDARY
3
CONNECT (4 - 5)
CONNECT (6 - 10)
CONNECT (11 - 12)
13



P1: (1 - 2), 4 Layers, 150 Turns per layer
(Total: 600 Turns), Cul 0.3mm
Insulating Foil 0.05mm between layers

i i : Insulating Foil 0.4mm

S1add: (3 - 4), 1 Layer 14 Turns 2 x Cul 1.8mm

i: Insulating Foil 0.05mm

S1: (5 - 6), 1 Layer 36 Turns Cul 1.4mm

i i : Insulating Foil 0.4mm

P2: (7 - 8 - 9), 10 Layers (5 + 5), 150 Turns per layer
(Total: 1500 (750 + 750) Turns), Cul 0.3mm
(5 Layers, TAP(8), 5 Layers)
Insulating Foil 0.05mm between layers

i i : Insulating Foil 0.4mm

S2: (10 - 11), 1 Layer 36 Turns Cul 1.4mm

i: Insulating Foil 0.05mm

S2add: (12 - 13), 1 Layer 14 Turns 2 x Cul 1.8mm

i i : Insulating Foil 0.4mm

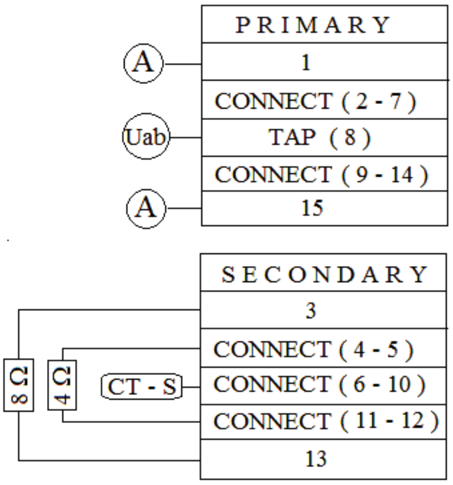
P2 / 2: (9 - 10), 5 Layers, 150 Turns per layer
(Total: 750 Turns), Cul 0.3mm
Insulating Foil 0.05mm between layers

i i : Insulating Foil 0.4mm

S2: (11 - 12), 1 Layer 36 Turns Cul 1.4mm

i i : Insulating Foil 0.4mm

P3: (14 - 15), 4 Layers, 150 Turns per layer
(Total: 600 Turns), Cul 0.3mm
Insulating Foil 0.05mm between layers



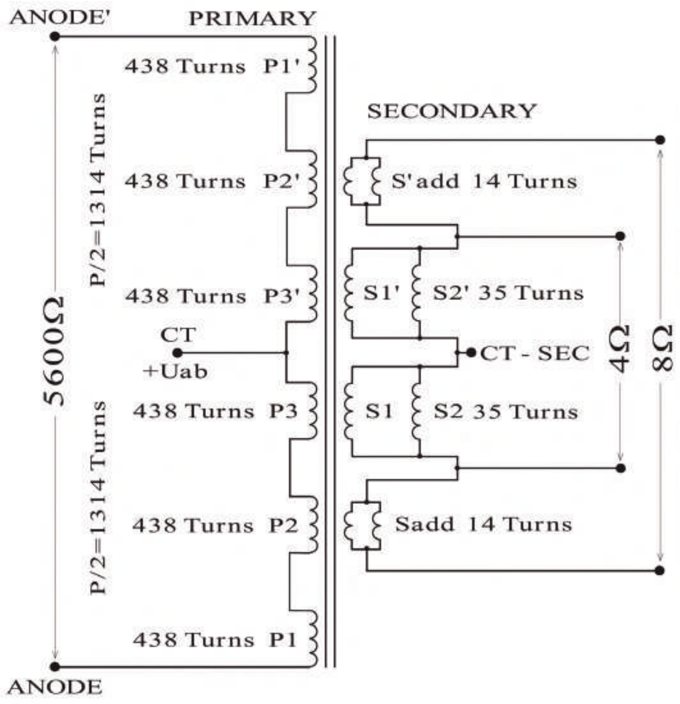
Push Pull Output Transformer:

$P_{OUT} = 20$ W
$R_{(a\text{-}a)} = 5600\ \Omega$
$R_L = 4\Omega, 8\Omega$
 # For example, it can be used in PP output stage with 2 x WE 300B ($U_{ak} = 350V$, $U_{gk} = 71V$, $I_a = 80mA$, $R_i = 720\ \Omega$)

Coil Former: (4 cm x 4 cm)
EI 120 4 % Si Fe transformer laminate, $\Delta = 0.35$ mm
$S_{Fe\ eff} = 15.36\ cm^2$

Number of turns of the Primary winding: 2700 Turns Cul Ø 0.3 mm
Number of turns of the 4Ω Secondary winding: 72 Turns Cul Ø 1.4 mm
Number of turns of the 8Ω Secondary winding: 101 (100) Turns
Inductance of the Primary winding: $L_P = 63\ H$
DC resistance of the Primary winding: $R_{P-DC} = 145.7\ \Omega$
Leakage Inductance: 11 mH
$f_{C0} = 14.15\ Hz$
$f_{sat} = 11.4\ Hz$
$f_{Low\ (-3dB)} = 3.08\ Hz$

Push Pull Output Transformer - <u>Improved AC and DC Symmetry</u>



Example (cont.)

Push Pull Output transformer
$P_{OUT} = 20\ W$
$R_{(a\text{-}a)} = 5600\ \Omega$
$R_L = 4\Omega, 8\Omega$

Transformer laminate EI 120 Fe 4 % Si Δ = 0.35mm
Coil former (4 x 4) cm split into two equal parts
Transformer effective cross-section
$S_{Fe\ eff} = 15.36\ cm^2$
$N_P = 2628$ *Turns, Cul Ø 0.3 mm*
$N_{S\ 4\Omega} = 70$ *Turns, Cul Ø 1.3 mm*
$N_{S\ 8\Omega} = 98$ *Turns*
(additional coil 14 Turns
2 x Cul Ø 1.7 mm, bifilar))

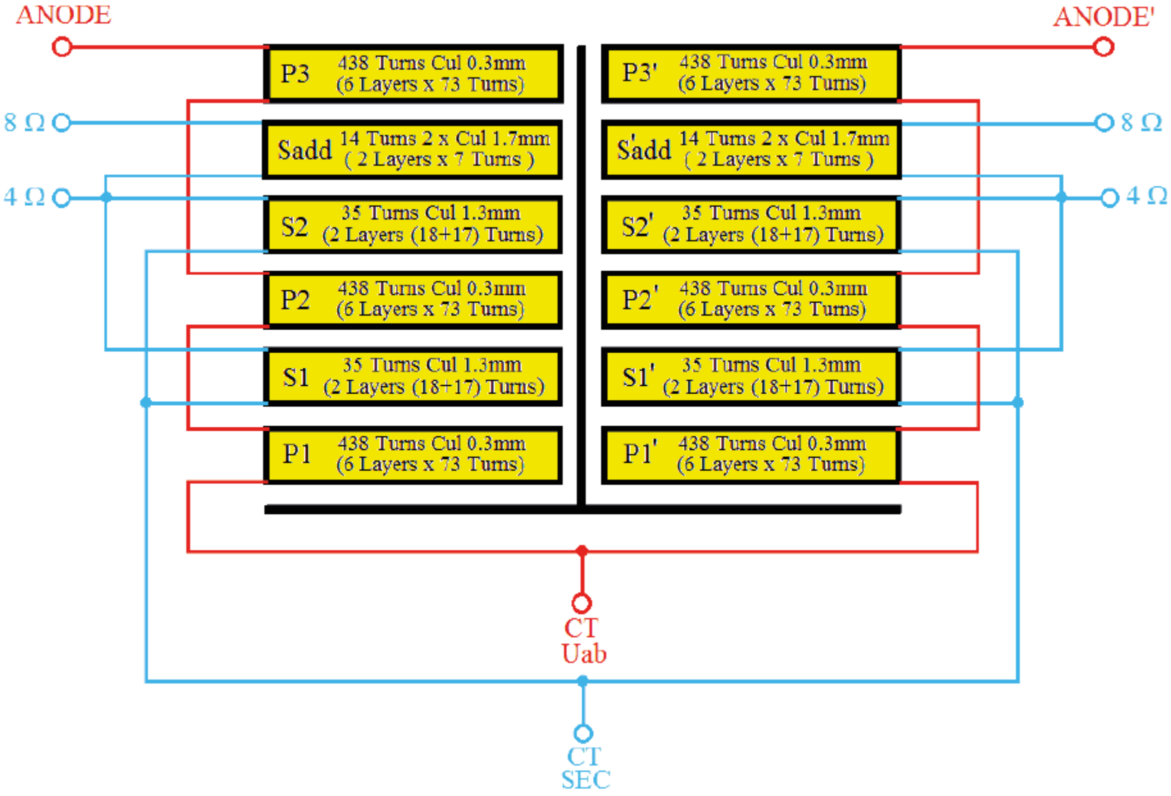
- *The Primary winding is split into two equal parts also:*
  $P / 2 = P' / 2 = \mathbf{N_P / 2} = 2628 / 2 = \mathbf{1314}$ ***Turns***
- *The each **half** of the Primary winding is split into **three** equal **sections connected in series**:*
  $P1 = P2 = P3\ (P1' = P2' = P3')$: $\mathbf{(N_P / 2) / 3 = 1314 / 3 = 438}$ ***Turns***
- *Each section of 438 turn is wound in **six layers** of 73 turns with wire Cul Ø 0.32 mm.*
- *The **beginnings** of the halves of the Primary winding (P / 2 and P' / 2) are **connected together** and this is the middle point of the Primary (CT –* U_{ab}*).*
- *The **ends** of the halves of the Primary winding are the connections for the **anodes** of the output tubes.*
- *The **Secondary** winding is split into **two equal parts connected** in series:*
  $N_{S\ 4\Omega} / 2 = 70 / 2 = 35$ Turns
- *Each of the two halves of the secondary is wound in **two** sections connected in parallel. Each section of 35 turns is wound in two layers with wire Cul Ø* 1.3 mm*: (18 + 17) turns = 35 turns.*
- The beginnings of the halves of the Secondary winding are connected together (CT – Sec).
- *The **ends** of the halves of the Secondary winding are the **connections** for the **4 Ω** load.*

- An additional winding is connected in series with each half of the 4Ω Secondary. Each additional winding of 14 turns is wound in two layers with two wires Cul Ø 1.7 mm (bifilar: parallel-winding, parallel connection):
  (7 + 7) turns = 14 turns.

- *The **ends** of the halves of the additional winding are the **connections** for the **8 Ω** load.*

- Insulation foil between P layers: # 0.05 mm; Insulation foil between S layers: # 0.05 mm.

- Insulation foil # 0.4 mm is placed between Primary and Secondary sections.



Winding procedure

Winding starts with winding the first half of the transformer:

Winding is carried out on one half of the coil former.

- *Winding the first section of the Primary, P1: 438 turns wound in 6 layers (73 turns per layer).*
  Wire: Cul Ø 0.3 mm.
  Insert one layer of insulating foil # 0.05 mm between the layers of wire.
  Finish the section with # 0.4 mm insulating foil.
- *Winding continue with winding the first section of the Secondary, S1: 35 turns wound in 2 layers ((18 +17) turns). Wire: Cul Ǿ 1.3 mm.*
  Insert one layer of the insulating foil # 0.05mm between the layers of wire.
  Finish the section with # 0.4mm insulating foil.
- *Winding the second section of the Primary, P2: 438 turns wound in 6 layers (73 turns per layer).*
  Wire: Cul Ø 0.3 mm.
  Insert one layer of insulating foil # 0.05 mm between the layers of wire.
  Finish the section with # 0.4 mm insulating foil.
- *Winding the second section of the Secondary, S2: 35 turns wound in 2 layers ((18 +17) turns). Wire: Cul Ǿ 1.3 mm.*
  Insert one layer of the insulating foil # 0.05mm between the layers of wire.
  Finish the section with # 0.4 mm insulating foil.

- *Winding the additional winding, S_{add} : 14 turns wound in 2 layers (7 turns per layer).*
  Wire: TWO wires Cul Ø 1.7 mm (bifilar: parallel-winding, parallel connection).
  Insert one layer of the insulating foil # 0.05mm between the layers of wire.
  Finish the section with # 0.4 mm insulating foil.
- *Winding the third section of the Primary, P3: 438 turns wound in 6 layers (73 turns per layer).*
  Wire: Cul Ø 0.3 mm.
  Insert one layer of insulating foil # 0.05 mm between the layers of wire.
  Finish the section with # 0.4 mm insulating foil.
- *Connect the Secondary sections S1 and S2 in parallel:*
  Connect all the beginnings of the Secondary sections together.
  Connect all the ends of the Secondary sections together.
  (Parallel connected Secondary sections S1 and S2 form half of the 4 Ω Secondary winding).
- *Connect the beginning of the additional winding to the end of the half of the 4 Ω Secondary winding explained above.*
- *Connect the end of the first Primary section to the beginning of the second Primary section.*
  *Connect the end of the second Pri*mary section to the beginning of the third Primary section.

Rotate the coil former 180^0 and wind the second half of the transformer repeating the procedure of winding the first half of the transformer explained above.
*(*The windings and sections of the second half of the transformer are denoted by the suffix " **'** ").*

- *Connect the **beginning** of the **first half of the Primary** (beginning of the section P1) and the **beginning** of the **second half** of the **Primary** (beginning of the section P1') – this is the **middle point** of the push-pull output transformer (CT – U_{ab}).*

- *Connect the **beginning** of the **first half** of the **4 Ω Secondary** (connected beginnings of the sections S1 and S2) and the **beginning** of the **second half** of the **4 Ω Secondary** (connected beginnings of the sections S1' and S2').*

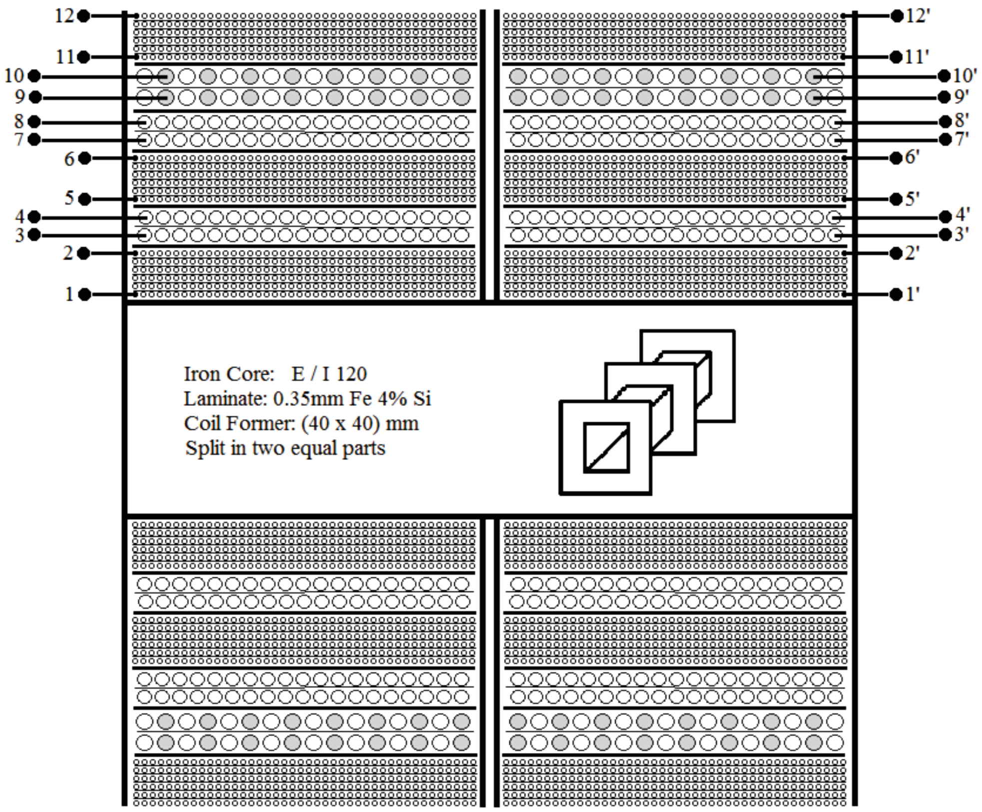
- *The **ends** of the halves of the **4 Ω Secondary** are the connections for the **4 Ω load** (loudspeaker).*

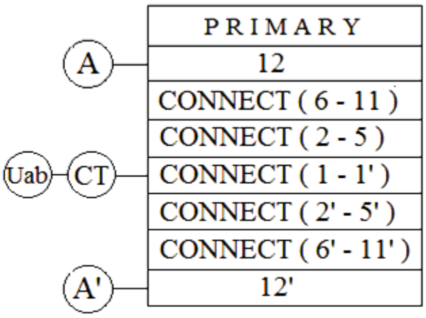
- *The **ends** of the **additional windings** (ends of the sections S_{add} and S'_{add} **)** are the connections for the **8 Ω load** (loudspeaker).*

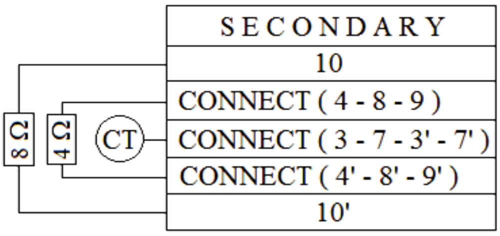
- ***EI 120** 4 % Si Fe transformer laminate. Laminate thickness: 0.35 mm.*

- *Stack the **E** laminates in the opposite direction piece by piece.*

- *Insert **I** laminates between E laminates.*


12
11
10
9
8
7
6
5
4
3
2
1
12'
11'
10'
9'
8'
7'
6'
5'
4'
3'
2'
1'
Iron Core: E / I 120
Laminate: 0.35mm Fe 4% Si
Coil Former: (40 x 40) mm
Split in two equal parts


PRIMARY
A
12
CONNECT (6 - 11)
CONNECT (2 - 5)
Uab
CT
CONNECT (1 - 1')
CONNECT (2' - 5')
CONNECT (6' - 11')
A'
12'


SECONDARY
10
CONNECT (4 - 8 - 9)
8 Ω
4 Ω
CT
CONNECT (3 - 7 - 3' - 7')
CONNECT (4' - 8' - 9')
10'

	P1 (1 – 2): 6 Layers, 73 Turns per layer (Total: 438 Turns) Cul Ø 0.3mm Insulating Foil 0.05mm between layers	**Calculated and expected:** *Push Pull Output Transformer (PP OT)* $P_{OUT} = 20W$ $R_{(a\text{-}a)} = 5600\Omega$ $R_L = 4\Omega, 8\Omega$ *Coil Former: (4cm x 4cm)* *EI 120 4 % Si Fe transformer laminate,* $\Delta = 0.35$ mm $S_{Fe\ eff} = 15.36$ cm^2 *Number of turns of the Primary:* *2628 turns . Cul Ø 0.3mm* *Number of turns of the 4Ω Secondary:* *70 turns. Cul Ø 1.3mm* *Number of turns of the 8Ω Secondary:* *98 Turns* *Inductance of the Primary:* $L_P = 59.68$ H *DC resistance of the Primary winding:* $R_{P-DC} = 138.8\ \Omega$ $f_{C0} = 14.9$ Hz $f_{sat} = 11.7$ Hz $f_{Low\ (-3dB)} = 3.2$ Hz
	ii: Insulating Foil 0.4mm	
	S1 (3 – 4): 2 Layers, (18 + 17) Turns (Total 35 Turns) Cul Ø 1.3mm Insulating Foil 0.05mm between layers	
	ii: Insulating Foil 0.4mm	
	P2 (5 – 6): 6 Layers, 73 Turns per layer (Total: 438 Turns) Cul Ø 0.3mm Insulating Foil 0.05mm between layers	
	ii: Insulating Foil 0.4mm	
	S2 (7 – 8): 2 Layers, (18 + 17) Turns (Total 35 Turns) Cul Ø 1.3mm Insulating Foil 0.05mm between layers	
	Sadd (9 – 10): 2 Layers, 7 Turns per layer (Total: 14 Turns) 2 x Cul Ø 1.7mm (bifilar) Insulating Foil 0.05mm between layers	
	ii: Insulating Foil 0.4mm	
	P3 (11 – 12): 6 Layers, 73 Turns per layer (Total: 438 Turns) Cul Ø 0.3mm Insulating Foil 0.05mm between layers	
The second half of the transformer (‘) is IDENTICAL to the first half		

 # Modified versions of the above transformer

1. $P_{OUT} = 20$ W, $R_{(a\text{-}a)} = 5600\ \Omega$,
   $\boldsymbol{R_L = 4\ \Omega}$

   The design of the transformer is similar to the above and differs in that the additional windings (S_{add} *and* S'_{add}) *are not wound.*

2. $P_{OUT} = 20$ W, $R_{(a\text{-}a)} = 5600\ \Omega$,
   $\boldsymbol{R_L = 8\ \Omega}$

   The design of the transformer is similar to the above and differs in that the additional windings (S_{add} *and* S'_{add} *) are not wound and instead of four sections of the 4 Ω Secondary (35 Turns in two layers), four sections of the 8 Ω Secondary are wound. Each section: 49 turns wound in two layers with wire Cul 0.95mm. Connecting the sections of the 8 Ω Secondary is identical to connecting the sections of the 4 Ω Secondary explained above.*

The purpose of the text above is to illustrate that there are many ways of winding techniques for PP transformers. The choice of winding arrangement and general construction of the PP output transformer depends on the design requirements and the technical knowledge and skill of the output transformer designer.

Ultra Linear PUSH PULL Output transformer

Some of the basic technical requirements that must be considered when designing a UL PP transformer

- ***The inductive coupling of the screen grid (g2) and the anode of the same tube has to be tighter than the inductive coupling of the screen grid (g2) and the anode of the other tube of the output stage.***

- ***The capacitive coupling of the screen grid (g2) of one tube and the anode of the other tube must be as low as possible.***

- ***The leakage inductance between the anode and screen grid of the same tube must be as low as possible.***

- ***The capacitances between the anodes and screen grids to the earth must be as low as possible.***

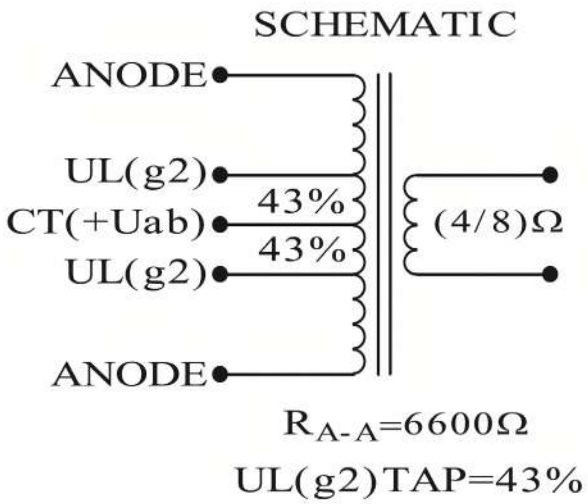
Output Transformer
Type: Ultra Linear PUSH PULL
R_P = 6600 Ω
Ultra Linear Tap: 43 %
R_S = 4 Ω, 8 Ω
Pout = (35 ÷ 40) W

*Iron Core**: EI 120 # 0.35 mm Fe 4% Si laminate***
*Coil Former**: (40 x 50) mm split in two equal parts***
Number of turns of the Primary, **N_P = 2760**
Number of turns of the 4 Ω Secondary, **$N_{S4\Omega}$ = 68**
Number of turns of the 8 Ω Secondary, **$N_{S8\Omega}$ = 96,**

Winding arrangement

P_1 *(6 layers x 77 Turns =* **462** *Turns* **Cul Ø 0.265 mm** *)* – **S_1** *(2 layers: 17 + 17 =* **34** *Turns* **Cul Ø 1.32 mm** *)* – **P_2** *(2 layers x 66 Turns =* **132** *Turns* **Cul Ø 0.315 mm** – **UL Tap** *+ 4 layers x 81 Turns =* **324** *Turns* **Cul Ø 0.265 mm***; Total:* **456** *Turns)* – **S_2** *(2 layers: 17 + 17 =* **34** *Turns* **Cul Ø 1.32 mm** *)* + **S_{add}** *(2 layer: 7 + 7 =* **14** *Turns* **2 x Cul Ø 1.7 mm** *(bifilar))* – **P_3** *(6 layers x 77 Turns =* **462** *Turns* **Cul Ø 0.265 mm** *)*

P_1' *(6 layers x 77 Turns =* **462** *Turns* **Cul Ø 0.265 mm** *)* – **S_1'** *(2 layers: 17 + 17 =* **34** *Turns* **Cul Ø 1.32 mm** *)* – **P_2'** *(2 layers x 66 Turns =* **132** *Turns* **Cul Ø 0.31 5mm** – **UL Tap** *+ 4 layers x 81 Turns =* **324** *Turns* **Cul Ø 0.26 5mm***; Total:* **456** *Turns)* – **S_2'** *(2 layers: 17 + 17 =* **34** *Turns* **Cul Ø 1.32mm** *) + S'_{add} (2 layer: 7 + 7 =* **14** *Turns* **2 x Cul Ø 1.7 mm** *(bifilar))* – **P_3'** *(6 layers x 77 Turns =* **462** *Turns* **Cul Ø 0.265 mm** *)*



Expected DC Resistance of the windings:

R_{DC-P} = 207 Ω; ($R_{DC-P/2}$ = 103.5 Ω)
$R_{DC-S4\Omega}$ = 0.0966 Ω

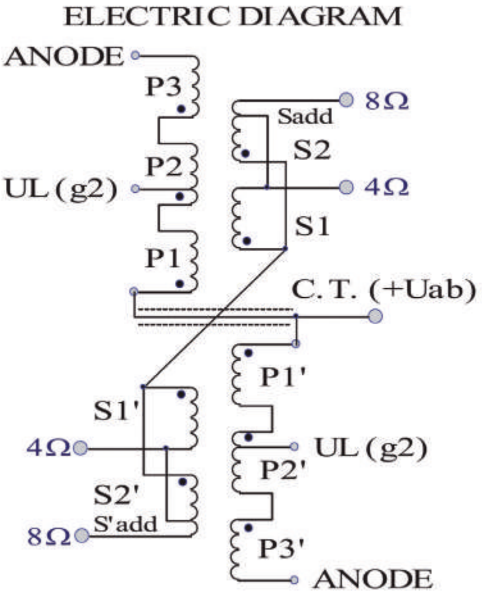
Cu loss:
$P_{P\ Cu\ loss}$ = 3.04 %
$P_{S4\Omega\ Cu\ loss}$ = 2.36 %
$P_{Total\ Cu\ loss}$ = **5.4 %**

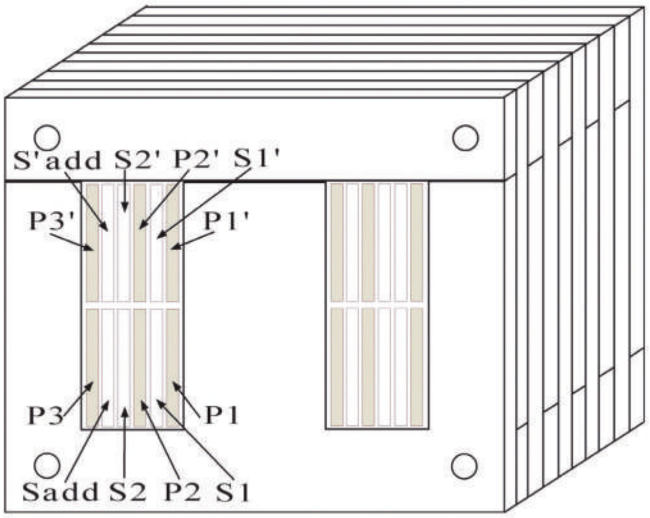
F_S = 12.5 Hz

F_{-3dB} = 4.2 Hz

L_P = 84 H

G_{Fe} = 3.3 kg





Winding procedure

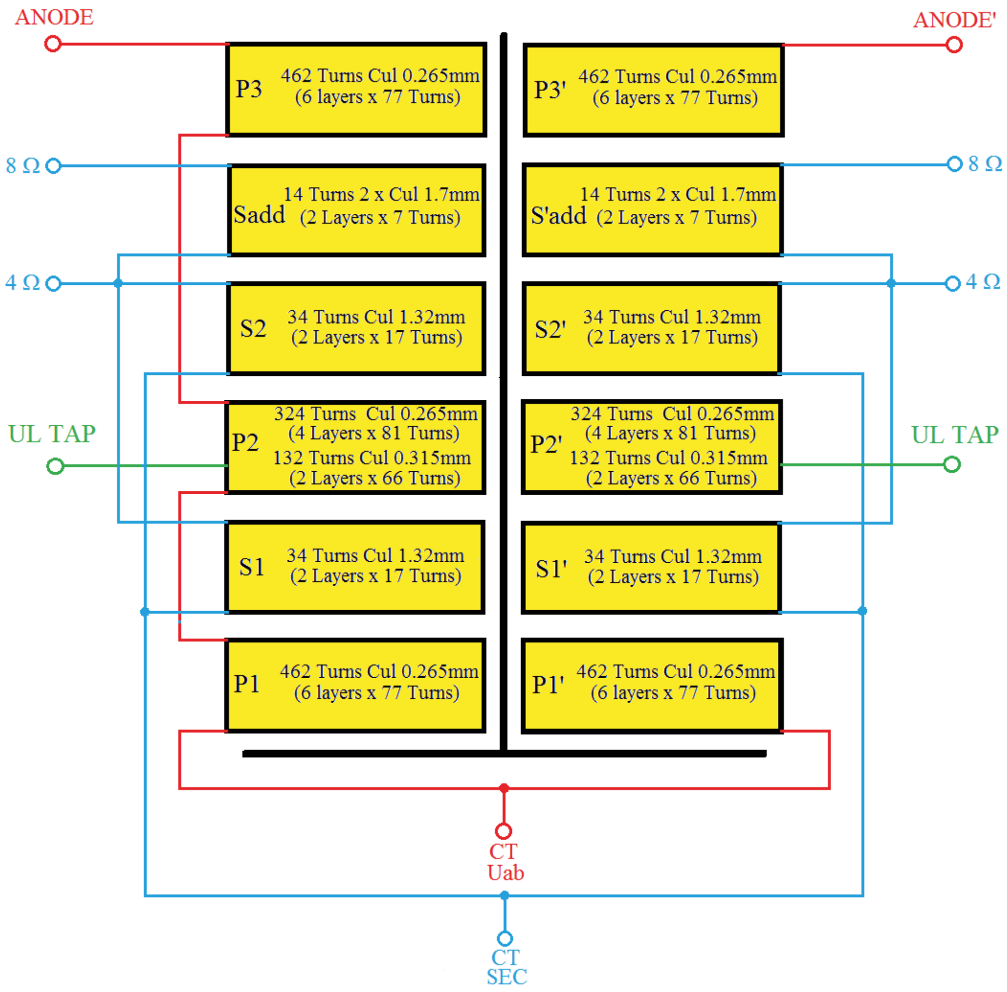
Winding starts with winding the first half of the transformer:
Winding is carried out on one half of the coil former.

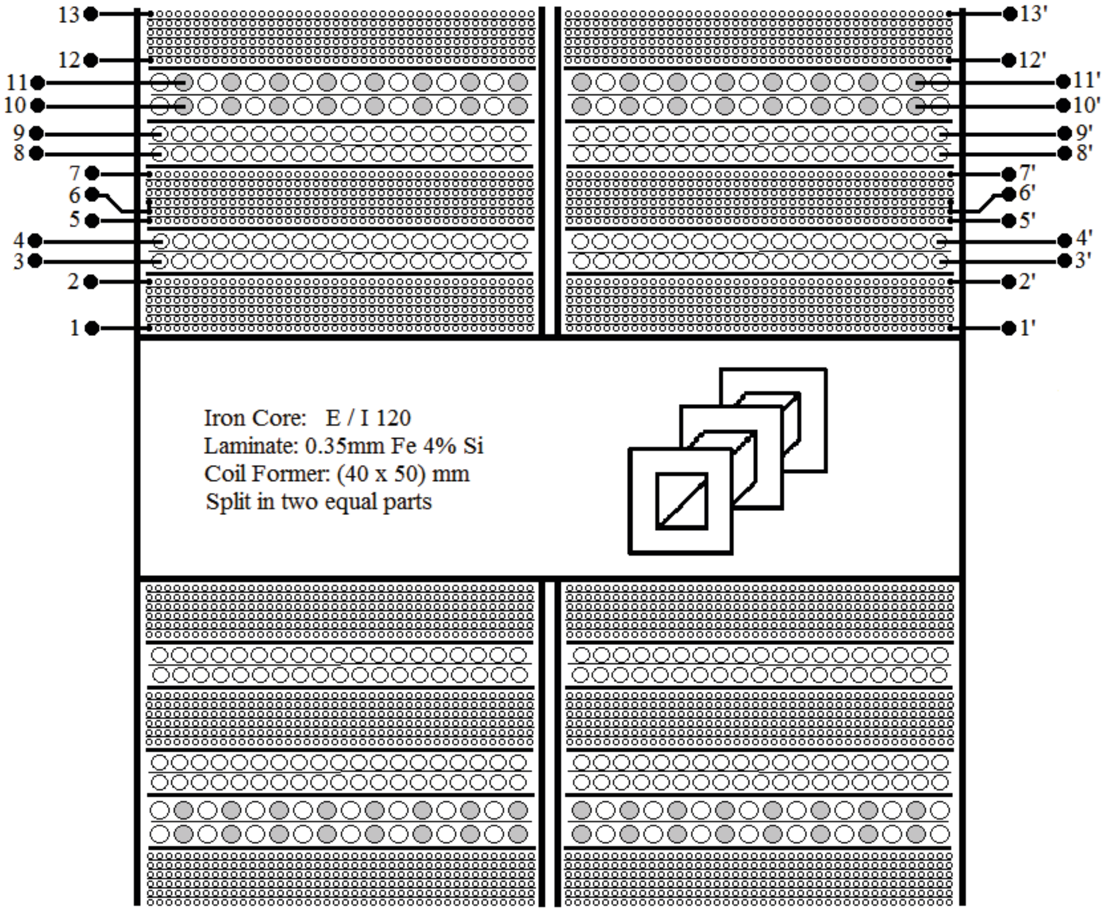
- *Winding the first section of the Primary (P1) with wire Cul Ø 0.265 mm. It has 462 turns wound in 6 layers of 77 turns each. (The beginning is marked with [1], the end of the winding is marked with [2]).*
  Insert one layer of insulating foil # 0.05 mm between the layers of wire.
  Finish the section with # 0.4 mm insulating foil.
- *Winding continues with winding the first section of the Secondary (S1) with wire Cul Ǿ 1.32 mm. It has 34 turns (17 +17) turns wound in two layers. (The beginning of the winding is marked with [3], the end of the winding is marked with [4]).*
  Insert one layer of the insulating foil # 0.05 mm between layers of wire.
  Finish the section with # 0.4mm insulating foil.
- *Winding the second section of the Primary (P2): 2 layers of 66 turns each (total: 132 turns), wire: Cul Ø 0.315 mm, make a TAP (UL) and continue winding 4 layers of 81 turns each (total 324 turns), wire: Cul Ø 0.265 mm. (The beginning of the winding is marked with [5], Tap is marked with [6], the end of the winding is marked with [7]).*
  Insert one layer of the insulating foil # 0.05 mm between layers of wire.
  Finish the section with # 0.4mm insulating foil.
- *Winding the second section of the Secondary (S2) with wire Cul Ǿ 1.3 mm. It has 34 turns (17 +17) turns wound in two layers. (The beginning of the winding is marked with [8], the end of the winding is marked with [9]).*
  Insert one layer of the insulating foil # 0.05mm between layers of wire.
- *Winding additional winding (S_{add}) with TWO wire Cul Ø 1.7 mm (bifilar). It has 14 turns (7 + 7) turns wound in two layers. (The beginning of the winding is marked with [10], the end of the winding is marked with [11]).*
  Insert one layer of the insulating foil # 0.05 mm between layers of wire.
  Finish the section with # 0.4 mm insulating foil.
- *Winding the third section of the Primary (P3) with wire Cul Ø 0.265 mm. It has 462 turns wound in 6 layers of 77 turns each. The beginning of the winding is marked with [12], the end of the winding is marked with [13]).*
  Insert one layer of the insulating foil # 0.05 mm between the wire layers.
- *Connect the beginning of the Secondary section S1 [3] and the beginning of the Secondary section S2 [8].*
- *Connect the end of the Secondary section S1 [4] and the end of the Secondary section S2 lead [9].*
- *Connect the beginning of the additional section S_{add} [10] and the ends of the Secondary sections S1 and S2, [4] and [9].*
- *Connect the end of the first Primary section (P1) [2] with beginning of the second Primary section (P2) [5]. Connect the end of the second Primary section (P2) [7] with beginning of the third Primary section (P3) [12].*

Rotate the coil former 180^0 and wind the second half of the transformer repeating the procedure of winding the first half of the transformer explained above.
*(*The windings and sections of the second half of the transformer are denoted by the suffix “ ‘ “).*

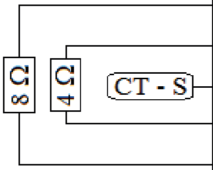
- *Connect the **beginning** of the **first half of the Primary** (beginning of the section P1) [1] and the **beginning** of the **second half** of the **Primary** (beginning of the section P1’) [1’] – this is the **middle point** of the push-pull output transformer (CT – U_{ab}).*
- *The **ends** of the halves of the **Primary (**end of the section P3 [13] and end of the section P3’ [13’]) are the connections for the anodes of the output tubes.*
- *The UL TAPs ([6] and [6’]) are the connections for the screen grids of the output tubes.*
- *Connect the **beginning** of the **first half** of the **4 Ω Secondary** (connected beginnings of the sections S1 and S2) [3] and [8] and the **beginning** of the **second half** of the **4 Ω Secondary** (connected beginnings of the sections S1’ and S2’) [3’] and [8’].*
- *The **ends** of the halves of the **4 Ω Secondary** (connected [4] and [9] and connected [4’[and [9’]) are the **4 Ω** (loudspeaker) load connections.*
- *The **ends** of the **additional windings** (ends of the sections S_{add} and S'_{add} **)** [11) and [11’] are the **8 Ω** (loudspeaker) load connections.*

- ***EI 120** 4 % Si Fe transformer laminate. Laminate thickness: 0.35 mm.*
- *Stack the **E** laminates in the opposite direction piece by piece.*
- *Insert **I** laminates between E laminates.*





	PRIMARY
A	13
	CONNECT (7 - 12)
UL	6
	CONNECT (2 - 5)
CT	CONNECT (1 - 1')
	CONNECT (2' - 5')
UL'	6'
	CONNECT (7' - 12')
A'	13'



SECONDARY
11
CONNECT (4 - 9 - 10)
CONNECT (3 - 8 - 3' - 8')
CONNECT (4' - 9' - 10')
11'

	P1 (1 – 2): 6 Layers, 77 Turns per layer (Total: 462 Turns) Cul Ø 0.265mm Insulating Foil 0.05mm between layers	***Calculated and Expected:*** *Ultra linear Push Pull Output transformer* $P_{OUT} = (35 \ldots 40)\ W$ $R_{(a-a)} = 6600\ \Omega$ UL Tap = 43 % $R_L = 4\ \Omega,\ 8\ \Omega$ *# Can be used in UL PP amplifier with 2 x KT88, 2 x EL34* *Coil Former: (4cm x 5cm)* *split in two equal parts* *EI 120 4 % Si Fe transformer laminate,* $\Delta = 0.35$ *mm* $S_{Fe\ eff} = 19.2\ cm^2$ *Number of turns of the Primary:* *2760 Turns Cul Ø 0.265mm* *Number of turns of the 4 Ω Secondary:* *68 Turns Cul Ø 1.32mm* *Number of turns of the 8 Ω Secondary:* *96 Turns* *Inductance of the Primary:* $L_P = 82.45\ H$ *DC resistance of the Primary winding:* $R_{P\ dc} = 204\ \Omega$ *Total Cu loss: 5.5 %* $f_{C0} = 12.75\ Hz$ $f_{sat} = 12.8\ Hz$ $f_{Low\ (-3dB)} = 4.96\ Hz$
	ii: Insulating Foil 0.4mm	
	S1 (3 – 4): 2 Layers, 17 Turns per layer (Total: 34 Turns) Cul Ø 1.32mm Insulating Foil 0.05mm between layers	
	ii: Insulating Foil 0.4mm	
	P2 (5 – 6 - 7): 2 Layers, 66 Turns per layer (Total: 132 Turns) Cul Ø 0.315mm Make UL Tap [6] Continue winding 4 Layers, 81 Turns per layer (Total: 324 Turns) Cul Ø 0.265mm Insulating Foil 0.05mm between layers	
	ii: Insulating Foil 0.4mm	
	S2 (8 – 9): 2 Layers, 17 Turns per layer (Total 34 Turns) Cul Ø 1.32mm Insulating Foil 0.05mm between layers	
	Sadd (10 – 11): 2 Layers, 7 Turns per layer (Total: 14 Turns) 2 x Cul Ø 1.7mm Insulating Foil 0.05mm between layers	
	ii: Insulating Foil 0.4mm	
	P3 (12 – 13): 6 Layers, 77 Turns per layer (Total: 462 Turns) Cul Ø 0.265mm Insulating Foil 0.05mm between layers	
The second half of the transformer (‘) is IDENTICAL to the first half		

A few more examples of the most commonly used Push Pull Output Transformers

I. *Output Transformer*
Type: Ultra Linear PUSH PULL
$R_P = 8000\ \Omega$
Ultra Linear Tap: 43 %
$R_S = 4\ \Omega,\ 8\ \Omega$
Pout = (35 … 40) W
 # For example, it can be used in PP output stage with 2 x KT88

Iron Core: EI 120 # 0.35 mm Fe 4 % Si laminate
Coil Former: (40 x 50) mm split in two equal parts

Number of turns of the Primary, N_P = 2792
Number of turns of the Secondary, $N_{S4\Omega}$ = 62
Number of turns of the Secondary, $N_{S8\Omega}$ = 88

Winding arrangement:

P_1 *(6 layers x 79 Turns =* ***<u>474</u>*** *Turns Cul Ø 0.265 mm)* – ***S_1*** *(2 layers: 15 + 16 =* ***31*** *Turns Cul Ø 1.32 mm)* – ***P_2*** *(2 layers x 66 Turns =* ***<u>132</u>*** *Turns Cul Ø 0.315 mm* – ***UL Tap*** *+ 4 layers x 79 Turns =* ***<u>316</u>*** *Turns Cul Ø 0.265 mm; Total:* ***<u>448</u>*** *Turns)* – ***S_2*** *(2 layers: 15 + 16 =* ***31*** *Turns Cul Ø 1.32 mm)* + ***S_{add}*** *(2 layers: 6 + 7 =* ***13*** *Turns 2 x Cul Ø 1.7 mm (bifilar))* – ***P_3*** *(6 layers x 79 Turns =* ***<u>474</u>*** *Turns Cul Ø 0.265 mm)*

P_1' *(6 layers x 79 Turns =* ***<u>474</u>*** *Turns Cul Ø 0.265 mm)* – ***S_1'*** *(2 layers: 15 + 16 =* ***31*** *Turns Cul Ø 1.32 mm)* – ***P_2'*** *(2 layers x 66 Turns =* ***<u>132</u>*** *Turns Cul Ø 0.315 mm* – ***UL Tap*** *+ 4 layers x 79 Turns =* ***<u>316</u>*** *Turns Cul Ø 0.265 mm; Total:* ***<u>448</u>*** *Turns)* – ***S_2'*** *(2 layers: 15 + 16 =* ***31*** *Turns Cul Ø 1.32mm)* + ***S_{add}*** *(2 layers: 6 + 7 =* ***13*** *Turns 2 x Cul Ø 1.7 mm (bifilar))* – ***P_3'*** *(6 layers x 79 Turns =* ***<u>474</u>*** *Turns Cul Ø 0.265 mm)*

Winding arrangement and winding procedure: as for UL PP 6600Ω transformer explained above.

Inductance of the Primary: L_P = 84.73 H
DC resistance of the Primary winding: R_{P-DC} = 206.7 Ω
Total Cu loss: 4.8 %

f_{C0} = 15 Hz
f_{sat} = 13.94 Hz
$f_{Low\ (-3dB)}$ = 5 Hz

II. *Output Transformer*
Type: Ultra Linear PUSH PULL
R_P = 4000 Ω
Ultra Linear Tap: 43 %
R_S = 4Ω, 8 Ω
Pout = 25 W
 # For example, it can be used in PP output stage with 4 x EL84 (6BQ5)

Iron Core: EI 120 # 0.35 mm Fe 4 % Si laminate
Coil Former: (40 x 40) mm split in two equal parts

Number of turns of the Primary, N_P = 2268
Number of turns of the Secondary, $N_{S4\Omega}$ = 72
Number of turns of the Secondary, $N_{S8\Omega}$ = 102

Winding arrangement:
P_1 *(6 layers x 63 Turns =* ***<u>378</u>*** *Turns Cul Ø 0.335 mm)* – ***S_1*** *(2 layers x 18 =* ***36*** *Turns Cul Ø 1.32 mm)* – ***P_2*** *(2 layers x 63 Turns =* ***<u>126</u>*** *Turns Cul Ø 0.335 mm* – ***UL Tap*** *+ 4 layers x 63 Turns =* ***<u>252</u>*** *Turns Cul Ø 0.335 mm; Total:* ***<u>378</u>*** *Turns)* – ***S_2*** *(2 layers x 18 =* ***36*** *Turns Cul Ø 1.32 mm)* + ***S_{add}*** *(2 layers: 7 + 8 =* ***15*** *Turns 2 x Cul Ø 1.5 mm (bifilar))* – ***P_3*** *(6 layers x 63 Turns =* ***<u>378</u>*** *Turns Cul Ø 0.335 mm)*
P_1' *(6 layers x 63 Turns =* ***<u>378</u>*** *Turns Cul Ø 0.335 mm)* – ***S_1'*** *(2 layers x 18 =* ***36*** *Turns Cul Ø 1.32 mm)* – ***P_2'*** *(2 layers x 63 Turns =* ***<u>126</u>*** *Turns Cul Ø 0.335 mm* – ***UL Tap*** *+ 4 layers x 63 Turns =* ***<u>252</u>*** *Turns Cul Ø 0.335 mm; Total:* ***<u>378</u>*** *Turns)* – ***S_2'*** *(2 layers x 18 =* ***36*** *Turns Cul Ø 1.32 mm)* + ***S_{add}*** *(2 layers: 7 + 8 =* ***15*** *Turns 2 x Cul Ø 1.5 mm (bifilar))* – ***P_3'*** *(6 layers x 63 Turns =* ***<u>378</u>*** *Turns Cul Ø 0.335mm)*

 # Insulation foil between layers # 0.05 mm
 # Insulation foil between section of the Primary and Secondary # 0.2 mm

Winding arrangement and winding procedure: as for UL PP 6600Ω transformer explained above.

Inductance of the Primary: L_P = 44.54 H
DC resistance of the Primary: R_{P-DC} = 98 Ω
Total Cu loss: 4.9 %

f_{C0} = 14.3 Hz
f_{sat} = 12.77 Hz
$f_{Low\ (-3dB)}$ = 4.9 Hz

III. *Output Transformer*
Type: Ultra Linear PUSH PULL
R_P = 8000 Ω
Ultra Linear Tap: 43 %
R_S = 4 Ω, 8 Ω
Pout = 15 W
 # For example, it can be used in PP output stage with 2 x EL84 (6BQ5)

Iron Core: EI 108 # 0.35 mm Fe 4 % Si laminate
Coil Former: (36 x 36) mm split in two equal parts

Number of turns of the Primary, N_P = 3312
Number of turns of the Secondary, $N_{S4\Omega}$ = 74
Number of turns of the Secondary, $N_{S8\Omega}$ = 104

Winding arrangement:
$\boldsymbol{P_1}$ *(6 layers x 92 Turns =* **552** *Turns Cul Ø 0.2 mm) –* $\boldsymbol{S_1}$ *(2 layers 18 + 19 =* **37** *Turns Cul Ø 1.06 mm) –*
$\boldsymbol{P_2}$ *(2 layers x 92 Turns =* **184** *Turns Cul Ø 0.2 mm –* ***UL Tap*** *+ 4 layers x 92 Turns =* **368** *Turns Cul Ø 0.2 mm;*
Total: **552** *Turns) –* $\boldsymbol{S_2}$ *(2 layers 18 + 19 =* **37** *Turns Cul Ø 1.06mm) +* $\boldsymbol{S_{add}}$ *(2 layers: 7 + 8 =* **15** *Turns*
2 x Cul Ø 1.25mm (bifilar)) – $\boldsymbol{P_3}$ *(6 layers x 92 Turns =* **552** *Turns Cul Ø 0.2 mm)*

$\boldsymbol{P_1'}$ *(6 layers x 92 Turns =* **552** *Turns Cul Ø 0.2 mm) –* $\boldsymbol{S_1'}$ *(2 layers 18 + 19 =* **37** *Turns Cul Ø 1.06 mm) –*
$\boldsymbol{P_2'}$ *(2 layers x 92 Turns =* **184** *Turns Cul Ø 0.2 mm –* ***UL Tap*** *+ 4 layers x 92 Turns =* **368** *Turns Cul Ø 0.2 mm;*
Total: **552***Turns) –* $\boldsymbol{S_2'}$ *(2 layers 18 + 19 =* **37** *Turns Cul Ø 1.06 mm) +* $\boldsymbol{S_{add}}$ *(2 layers: 7 + 8 =* **15** *Turns*
2 x Cul Ø 1.25mm (bifilar)) – $\boldsymbol{P_3'}$ *(6 layers x 92 Turns =* **552** *Turns Cul Ø 0.2 mm)*

 # Insulation foil between layers # 0.05 mm
 # Insulation foil between sections of the Primary and Secondary # 0.4 mm

Winding arrangement and winding procedure: as for UL PP 6600Ω transformer explained above.

Inductance of the Primary: L_P = 85.48 H
DC resistance of the Primary: $R_{P\ -\ DC}$ = 367 Ω
Total Cu loss: 7.8 %

f_{C0} = 14.9 Hz
f_{sat} = 11.87 Hz
$f_{Low\ (-3dB)}$ = 5.2 Hz

IV. *Output Transformer*
Type: Ultra Linear PUSH PULL
R_P = 3300 Ω
Ultra Linear Tap: 44 %
R_S = 4 Ω, 8 Ω
Pout = 70 W
 # For example, it can be used in PP output stage with 4 x KT88

Iron Core: EI 150 # 0.35 mm Fe 4 % Si laminate
Coil Former: (50 x 50) mm split in two equal parts

Number of turns of the Primary, $N_P = 1956$
Number of turns of the Secondary, $N_{S4\Omega} = 68$
Number of turns of the Secondary, $N_{S8\Omega} = 96$

Winding arrangement:
P_1 (6 layers x 54 Turns = **324** Turns Cul Ø 0.4 mm) **–** **S_1** **(**2 layers 34 + 34 = **68** Turns Cul Ø 0.85 mm) **–** **P_2** (2 layers x 54 Turns = **108** Turns Cul Ø 0.4 mm **– UL Tap** + 4 layers x 54 Turns = **216** Turns Cul Ø 0.4 mm; Total: **324** Turns) – **S_2** (2 layers 34 + 34 = **68** Turns Cul Ø 0.85 mm) **+** **S_{add}** (2 layers: 14 + 14 = **28** Turns Cul Ø 2.12 mm) – **P_3** (6 layers x 55 Turns = **330** Turns Cul Ø 0.4 mm)

P_1' (6 layers x 54 Turns = **324** Turns Cul Ø 0.4mm) – **S_1'** (2 layers 34 + 34 = **68** Turns Cul Ø 0.85mm) – **P_2'** (2 layers x 54 Turns = **108** Turns Cul Ø 0.4mm – **UL Tap** **+** 4 layers x 54 Turns = **216** Turns Cul Ø 0.4mm; Total: **324** Turns) – **S_2'** (2 layers 34 + 34 = **68** Turns Cul Ø 0.85mm) + **S'_{add}** (2 layer: 14 + 14 = **28** Turns Cul Ø 2.12mm) – **P_3'** (6 layers x 55 Turns = **330** Turns Cul Ø 0.4mm)

 # Insulation foil between layers # 0.05 mm
 # Insulation foil between sections of the Primary and Secondary # 0.4mm

Windings connection:
P1, P2, P3 - connected on series
P1', P2', P3' - series connected
Beginning of section P1 and beginning of section P1' - connected in series (CT, +Uab)
End of section P3 and end of section P3' – the anodes of the output tubes
S1, S2, S1', S2' – connected in parallel - 4Ω load
S_{add} and S'_{add} – connected in parallel - 8Ω load. The beginnings of sections S_{add} and S'_{add} connected to the ends of 4Ω sections S1, S2, S1', S2' .

Inductance of the Primary: $L_P = 41.4$ H
DC resistance of the Primary: $R_{P\text{ - }DC} = 76.34\ \Omega$
Cu loss: 5.8 %
$f_{C0} = 12.69$ Hz
$f_{sat} = 14.4$ Hz
$f_{Low\ (-3dB)} = 5.2$ Hz

V. *Output Transformer*
Type: Ultra Linear PUSH PULL
$R_P = 3300\ \Omega$
Ultra Linear Tap: 44 %
$R_S = (4 / 6)\ \Omega$, (8 / 10) Ω, 16 Ω
Pout = 30W

 # Construction of the UL PP output transformer for mass production, which is most often used by manufacturers.

Iron Core: EI 120 # 0.35 mm Fe 4 % Si laminate
Coil Former: (40 x 40) mm split in two equal parts
Number of turns of the Primary, $N_P = 2070$
Number of turns of the Secondary, $N_{S4\Omega} = 72$

Winding arrangement:
P_1 (3 layers x 58 Turns + 3 layers x 57 Turns = **345** Turns Cul Ø 0.375 mm) – **S_1** (2 layers 18 + 18 = **36** Turns Cul Ø 1.32 mm) – **P_2** (2 layers x 57 Turns = **114** Turns Cul Ø 0.375 mm – **UL Tap** + 1 layer x 57 Turns + 3 layers x 58 Turns = **231** Turns Cul Ø 0.375 mm ; Total: **345** Turns) – **S_2** (2 layers 18 + 18 = **36** Turns Cul Ø 1.32 mm) – **P_3** (3 layers x 58 Turns + 3 layers x 57 Turns = **345** Turns Cul Ø 0.375 mm)

P_1' (3 layers x 58 Turns + 3 layers x 57 Turns = **345** Turns Cul Ø 0.375 mm) – **S_1'** (2 layers 18 + 18 = **36**Turns Cul Ø 1.32 mm) – **P_2'** (2 layers x 57 Turns = **114** Turns Cul Ø 0.375 mm – **UL Tap** + 1 layer x 57 Turns + 3 layers x 58 Turns = 231 Turns Cul Ø 0.375 mm ; Total: **345** Turns) – **S_2'** (2 layers 18 + 18 = **36** Turns Cul Ø 1.32 mm) – **P_3'** (3 layers x 58 Turns + 3 layers x 57 Turns = **345** Turns Cul Ø 0.375 mm)

 # Insulation foil between layers # 0.05 mm
 # Insulation foil between sections of the Primary and Secondary # 0.4 mm

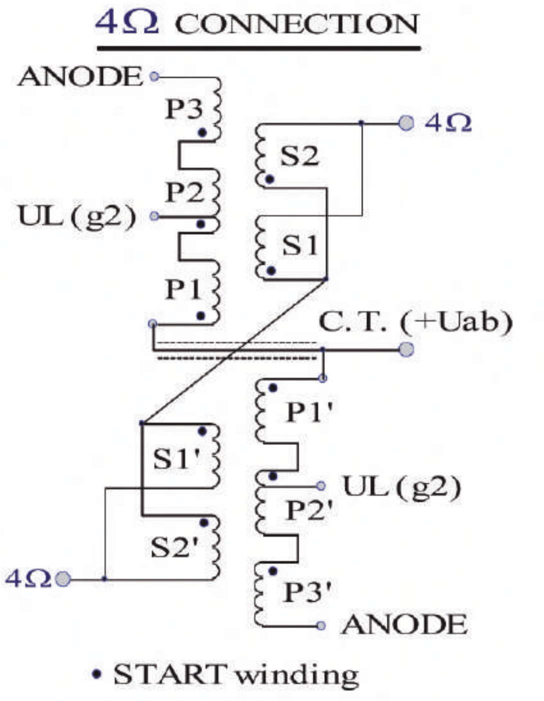
Windings connection:

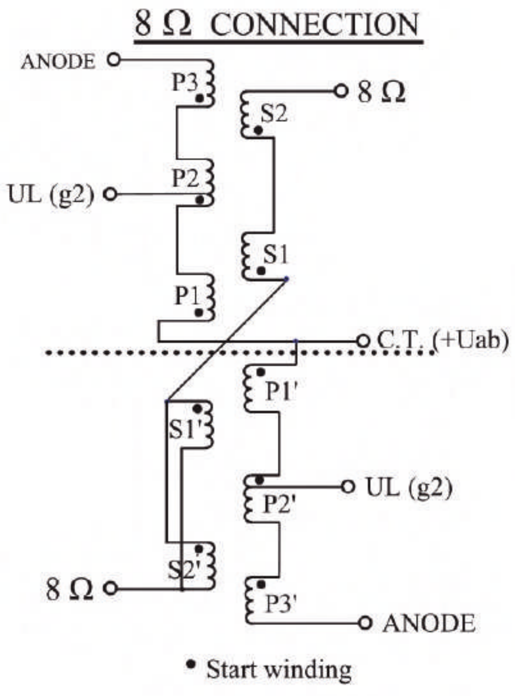
P1, P2, P3 – connected in series
P1', P2', P3' - connected in series
Beginning of section P1 and beginning of section P1' - connected in series (CT, +Uab)
End of section P3 and end of section P3' – the anodes of the output tubes

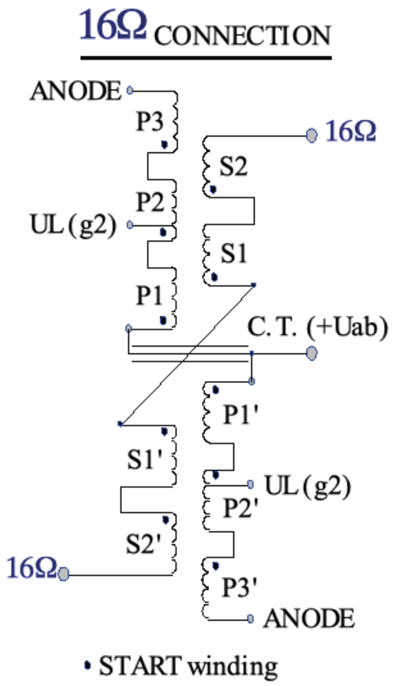
Inductance of the Primary: L_P *= 37.1 H*
DC resistance of the Primary: $R_{P\text{ - }DC}$ *= 71.8 Ω*
Cu loss: 4.5 %

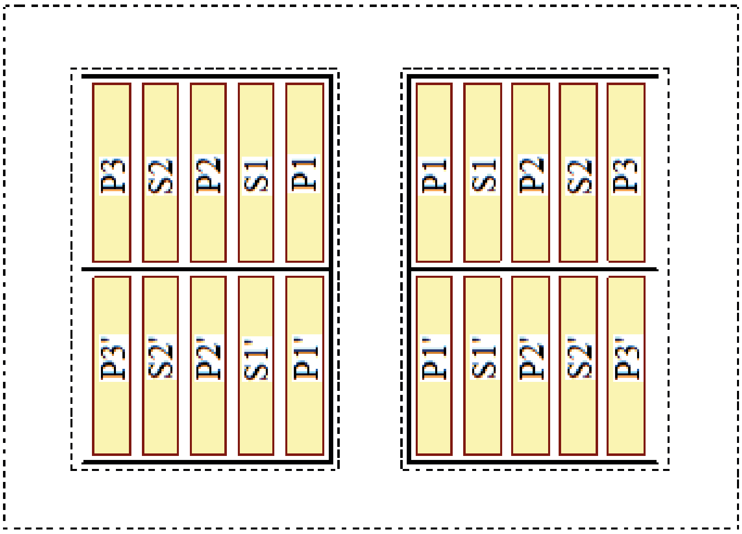
f_{C0} *= 14 Hz*
f_{sat} *= 13.9 Hz*
$f_{Low\ (-3dB)}$ *= 7.7 Hz*

Connecting the windings of the Secondary
It should be noted that each of the sections S1. S2, S1', S2' is actually ***1Ω Secondary****.*









PUSH PULL output transformer designing flow chart:

FIRST STEP

PROJECT REQUIREMENTS – INPUT DATA:

- The operating point of the output tube (Quiescent point).
- Internal resistance (R_i) of the output tube
- Plate (anode) to plate (anode) load (R_{La})
- Output power
- Frequency band
- Secondary load

SECOND STEP

First method:

- Calculating the cross- section of the iron core
- $S_{Fe}(cm^2) = \dfrac{1.8\times10^6 \times P(W)}{f(Hz)\times B(gauss)\times S_W(cm^2)\times J\left(\frac{A}{mm^2}\right)}$

B (Gauss) – Magnetic Flux Density
f (Hz) – The lowest frequency (Hz) at which the transformer can operate at full power
J (A/mm^2) – Electric Current Density
P (W) – Transformer power
S_W (cm^2) – Cross-section of the transformer core window (For E/I laminate S_W = (a x c) [cm^2]

A simplified equation (f = (18 ... 20) Hz. J = (1.5 ... 2) A / mm^2, B = 8000 Gauss):

$$S_{Fe}(cm^2) = \frac{(6.25 \div 8.3) \times P_P(W)}{S_W(cm^2)}$$

Second method:

$$S_{Fe(min)}(cm^2) = \sqrt{\frac{P \times G \times 10^6}{B \times J \times f}}$$

P **–** Output power [W]
B **–** Induction **(8000 Gs)** (4% Si Ferro alloy transformer laminates)
G **–** Coefficient of weight ratio of iron core and cooper winding **(1.7 … 2.5)**
J **–** Electric Current Density **(1.5 … 2.5) A / mm^2**
f **–** The lowest frequency (Hz) at which the transformer can operate at full power

A simplified equation (f = (18 ... 20) Hz. J = (1.5 ... 2) A / mm^2, B = 8000 Gauss):

$$S_{Fe(min)}(cm^2) = 2.95 \times \sqrt{P} \approx (3 - 3.4) \times \sqrt{P}$$

- Choosing a standard transformer laminate and coil former

- Calculating: S_{eff} , S_w , l_{Fe} , V_{Fe} , G_{Fe}
  $S_{Fe\ eff}$ = (0.92 – 0.96) x (b x h)
  S_W = a x c
  ℓ_{Fe} = 2 x c + 2 x a + π x a ℓ_{Fe} = 5.57 x b
  V_{Fe} = $S_{Fe\ eff}$ x l_{Fe}
  G_{Fe} = D_{Fe} x V_{Fe}

THIRD STEP

Calculating the number of turns of the Primary and Secondary windings (N_P and N_S)

$$L = \frac{R_i \times R_a}{R_i + R_a} \times \frac{1}{2 \times \pi \times f_{Low(-3dB)}}$$

$$L = \frac{1}{2 \times \pi \times f} \times \frac{R_i \times R_a}{R_i + R_a} \times \frac{1}{\sqrt{10^{0.1 \times b} - 1}}$$

A simplified equation for calculating the inductance of the Primary (R_{a-a} = 4 x R_i and ,$f_{Low(-3dB)}$ = 2.5Hz, or impedance of the Primary equal to the R_{a-a} at f = 15Hz):

$$L_P \geq \frac{R_{a-a}}{94.25}$$

Number of turns of the Primary:

$$\boldsymbol{N_P = 8923 \times \sqrt{\frac{2 \times L[H] \times l_{Fe}[cm]}{S_{eff}[cm^2] \times \mu_e}} \approx 9000 \times \sqrt{\frac{2 \times L[H] \times l_{Fe}[cm]}{S_{eff}[cm^2] \times \mu_e}}}$$

$$\boldsymbol{N_P = 8923 \times \sqrt{\frac{L_P[H] \times l_{Fe}[cm]}{S_{eff}[cm^2] \times \mu_e}} \approx 9000 \times \sqrt{\frac{L_P[H] \times l_{Fe}[cm]}{S_{eff}[cm^2] \times \mu_e}}}$$

μ_e ≈ **1000 (**as closely as possible stacked 4% Si Fe E and I laminates)

Class B Push Pull amplifier
Number of turns of the Primary

$$\boldsymbol{N_P \approx 9000 \times \sqrt{\frac{4 \times L[H] \times l_{Fe}[cm]}{S_{eff}[cm^2] \times \mu_e}}}$$

Number of turns of the Secondary:

$$N_S = N_P \times \sqrt{\frac{R_S}{R_{a-a}}}$$

N_P – Number of turns of the Primary
N_S – Number of turns of the Secondary
$R_{a\text{-}s}$ – Load resistance of the Primary (Ω)
R_S – Load resistance of the Secondary (Ω)

Wire diameter of the Primary and Secondary windings:

Wire diameter of the Primary winding:

$$d_{Pmin}(mm) = 1.13 \times \sqrt{\frac{I_a(A)}{J(\frac{A}{mm^2})}}$$

$I_{P(DC)}$ *–Plate (Anode) DC Current (**one tube**)*

$$For\ J = 1.5\frac{A}{mm^2}: \quad d_{Pmin}(mm) \approx 0.9 \times \sqrt{I_{P(DC)}(A)}$$

$$For\ J = 2\frac{A}{mm^2}: \quad d_{Pmin}(mm) \approx 0.8 \times \sqrt{I_{P(DC)}(A)}$$

$$For\ J = 2.5\frac{A}{mm^2}: \quad d_{Pmin}(mm) \approx 0.7 \times \sqrt{I_{P(DC)}(A)}$$

Choise of standard wire diameter: $\mathbf{d_{P(standard)} \geq d_{Pmin}}$

FOURTH STEP

- **Calculating the minimum wire diameter of the Primary and Secondary winding**

$$d_{Pmin}(mm) = 1.13 \times \sqrt{\frac{I_a(A)}{J(\frac{A}{mm^2})}} \qquad I_S = \frac{N_P}{N_S} \times \sqrt{\frac{P}{R_{La}}}$$

$$d_{Wp}(mm) = \sqrt{\frac{0.28 \times S_W(mm^2)}{N_P}} \quad \rightarrow \quad N_{P(max)} = \frac{0.28 \times S_W}{d_{Wp}^2}$$

$$Turn/_{Layer} = \frac{0.97 \times c''}{d_{Wp(standard)}} = \frac{0.97 \times (c - 4mm)}{d_{Wp(standard)}\ (mm)}$$

$$n_{layer} = \frac{N_P}{\left(\frac{Turns}{Layer}\right)}$$

FIFTH STEP

- **Choosing the winding arrangement**

SIXTH STEP

- **Recalculating the diameter of the wires and checking the possibility of placing all the windings inside the window of the transformer core.**

$$height_{total} = n_{Players} \times (d_{P\,with\,isolation} + \Delta_i) + n_{Slayers} \times (d_{S\,with\,isolation} + \Delta_i) + n_{P-S\,layers} \times \Delta_{ii}$$

- **Recalculating the number of turns of the Primary and Secondary windings.**

SEVENTH STEP

- **Calculating the resistance of the windings**

- **Calculating the Cu loss**

$$\ell_t = 2 \times \mathrm{h} + \pi \times \mathrm{h}_0 + 2 \times \mathrm{b} + 4 \times \mathrm{s} \qquad \ell_{wP} = N_P \times \ell_t$$

$$R_{c-dc} = \rho \times \frac{4 \times \ell_w}{\pi \times d^2} \qquad \boldsymbol{P}_{P-loss}(\%) = 100 \times \frac{R_{P(c-dc)}}{R_{P(c-dc)} + R_{La}}$$

$$\boldsymbol{P}_{S-loss}(\%) = 100 \times \frac{R_{S(c-dc)}}{R_{S(c-dc)} + R_S} \qquad \boldsymbol{P}_{loss}(\%) = \boldsymbol{P}_{P-loss}(\%) + \boldsymbol{P}_{S-loss}(\%)$$

EIGHT STEP

- Calculating L_p , L_s, f_{Low}, f_{-3dB}, f_{C0}, B_{dc}, B_{ac}, B_{tot}, f_{sat}

$$L_P = \frac{1.256 \times S_{eff} \times \mu_e \times N_P^2}{10^8 \times l_{Fe}} \qquad R_A = \frac{(2 \times R_i + R_{Pdc}) \times R_{La-a}}{2 \times R_i + R_{Pdc} + R_{La=a}}$$

$$f_{low} = \frac{R_A}{2 \times \pi \times L_P} \qquad f_{C_0} = \frac{R_{La-a}}{2 \times \pi \times L_p}$$

$$f_{sat} = \frac{22.6 \times U_{eff} \times 10^6}{S_{eff} \times N_p \times B_{max}} \qquad L_S = \frac{0.417 \times Np^2 \times l_t \times [(2 \times n_{P-S} \times \Delta_{ii}) + h_{total}]}{10^9 \times n_{P-D}^2 \times c''}$$

$$\boldsymbol{R_B} = R_i + R_{Pdc} + R_{La} \qquad f_{(-3dB)high} = \frac{R_B}{2\pi \times L_S}$$

NINTH STEP

- **Making, testing and measuring transformer characteristics**

5.7 POWER SUPPLY TRANSFORMERS

The power transformer supplies electronic circuits with the various voltages necessary for their operation.

Input data required for designing a power transformer:

- Mains voltage or Primary voltage (**U_p**) and frequency (**f**) of the mains.
- Voltage (**U_s**) and current (**I_s**).of the Secondary

Cross-section of the Iron core (E/I laminates)

$$S_{Fe(min)}(\boldsymbol{cm^2}) = \sqrt{\frac{P_{0maxac} \times G \times 10^6}{B \times J \times f}}$$

P_{0maxac} – Output power [W] of the Secondary
B – Induction **(Gauss)** (4% Si Ferro alloy transformer laminates)
G – Coefficient of weight ratio of iron core and cooper winding **(1.0 … 1.5)**
J – Electric Current Density **(1.5 … 2.5) A / mm^2**
f – frequency (Hz) of the mains

For J = 2.5 A/mm^2 , G = 1.5, B = 12000 Gs:

$$S_{Fe}(cm^2) = 7 \times \sqrt{\frac{P_S(VA)}{f(Hz)}} = 7 \times \sqrt{\frac{U_S(V) \times I_S(A)}{f(Hz)}} \quad (501.56)$$

f (Hz) – frequency of the mains
P_s – Total power of the Secondary

The above formula is used for the calculation of power transformers used in the mass production of consumer electronics and the reason is the lower production cost: less EI transformer laminate and less Cu wire.

If the transformer has more than one secondary, it is necessary to calculate the total power of all secondaries:

$$P_s = (U_{s1} \times I_{s1}) + (U_{s2} \times I_{s2}) + \cdots (U_{sn} \times I_{sn}) = \sum_{i=1}^{n} (U_{si} \times I_{si})$$

Transformer laminate:
Calculation of the optimal dimension of the transformer laminates (b):

$$b_{(max)}(cm) = \sqrt{S_{Fe}}$$

$$b_{(min)}(cm) = 0.85 \times \sqrt{S_{Fe}} = 0.85 \times b_{(max)}$$

$$b_{(min)} < b < b_{(max)}$$

Choose a standard transformer laminate

Calculation of the height of the stack of transformer laminates (h):

$$h(cm) = \frac{S_{Fe}(cm^2)}{k_i \times b(cm)}$$

k_i – stacking coefficient (0.92 – 0.98)

EI laminate Fe 4 % Si with a thickness of 0.5 mm is most often used for the production of power transformers.

k_i – 0.97 (EI laminate Δ = 0.5 mm)

 # Choose a standard coil former (or make it if it is not of standard dimensions).

N_P – <u>Number of turns of the Primary</u>

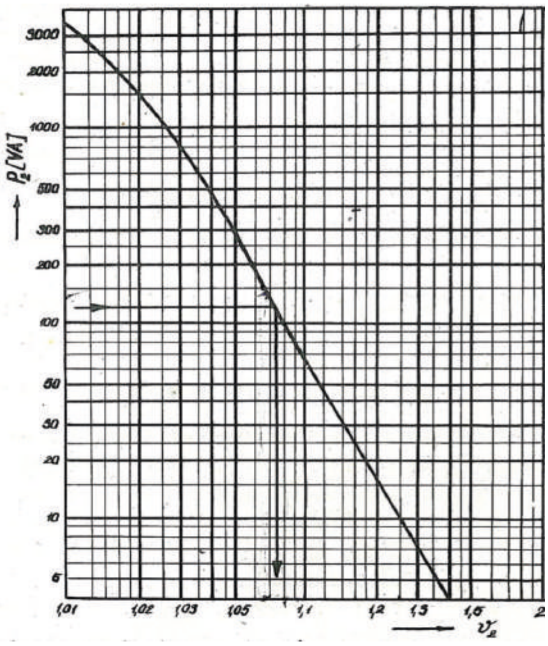
$$N_p = \frac{U_p \times 10^4}{4.44 \times f \times B_m \times S_{Fe}} \quad (501.57)$$

Turn per Volts (T / V):
A simplified equation (f = 50 Hz. B_m = 1T (10 000 Gauss):

$$\frac{Turn}{V} \approx \frac{45}{S_{Fe}} \quad (501.58)$$

and $$N_p = U_p \times \frac{Turn}{V} \approx U_p \times \frac{45}{S_{Fe}} \tag{501.59}$$

N_s – <u>Number of turns of the Secondary</u>



$$N_s = \frac{U_s(1+\frac{\Delta U}{U_s})\times 10^4}{4.44 \times f \times B_m \times S_{Fe}} \tag{501.60}$$

Coefficient of loss of power of the Secondary:

$$\mathbf{1} + \frac{\Delta \boldsymbol{U}}{\boldsymbol{U}_s} = \boldsymbol{v}_2;\ \boldsymbol{v}_2\ from\ the\ diagram\text{:}\ \boldsymbol{v}_2 = \boldsymbol{f}(\boldsymbol{P}_S)$$

or

$$N_s = \frac{U_s \times N_p}{U_p \times \sqrt{\eta}} \tag{501.61}$$

η – ***efficiency of the transformer core = 0.82 (0.8)***
(Fe Si 4 % NOSS Δ = 0.5 mm)

<u>Calculation of the wire diameter of the transformer winding</u>

The wire diameter of the Primary winding:

$$I_p = \frac{P_p}{U_p \times \sqrt{\eta}}$$

$$d_p = 1.13 \times \sqrt{\frac{I_p}{J}}$$

The wire diameter of the Secondary winding:

$$d_s = 1.13 \times \sqrt{\frac{I_s}{J}}$$

Core laminate material	**Magnetic flux density φ [Wb/m²]**	**Core efficiency η [1/1]**
Grain-oriented silicon steel (C-shaped), M5	1.3	0.88
Grain-oriented silicon steel (0.35 mm plates), M6	1.2	0.84
Non grain-oriented silicon medium steel (0.5 mm plates), M7	1.1	0.82
Non grain-oriented silicon standard steel (or for heavy duty)	1.0	0.80
Mild steel	0.8	0.70

Further calculation is similar to the transformer calculation already explained (adjusting N_P and N_S as an integer, number of layers, turns per layer, standard diameter of Cu wire according to the possibility of fitting the windings in the window of the transformer core, DC resistance of the Primary and Secondary windings, Cu loss, etc).

Example:
Mains voltage: 220V / 50Hz

Secondary:
S_1 - 2 x 6.3 V / 1.5 A
S_2 - 12.6 V / 1 A
S_3 - (240 + 110) = 350 V / 0.3 A

The total power of all secondaries:

$$P_s = (U_{s1} \times I_{s1}) + (U_{s2} \times I_{s2}) + (U_{s3} \times I_{s3})$$

$$P_s = 2 \times (6.3 \times 1.5) + (12.6 \times 1) + (350 \times 0.3) = 136.5\ W$$

Cross-section of the Iron core:

$$S_{Fe(min)}(cm^2) = \sqrt{\frac{P_{0_{maxac}} \times G \times 10^6}{B \times J \times f}}$$

From the table (NOSS transformer laminate): $B_m = 1T = 10000$ *Gauss and* **η** *= 0.8*

$$S_{Fe(min)}(cm^2) = \sqrt{\frac{P_{0_{maxac}}}{\eta} \times \frac{G \times 10^6}{B \times J \times f}} = \sqrt{\frac{136.5}{0.8} \times \frac{1.25 \times 10^6}{10000 \times 2.5 \times 50}} = 13\ cm^2$$

Core laminate:

$$b(cm) = \sqrt{S_{Fe}} = \sqrt{13} = 3.6$$

Standard transformer laminate (b = 3.6 cm): EI 108

$$h(cm) = \frac{S_{Fe}(cm^2)}{k_i \times b(cm)} = \frac{13}{0.97 \times 3.6} = 3.7\ cm$$

Standard coil former: (3.6 x 3.6) cm

$$S_{Fe_{eff}} = 0.97 \times 3.6 \times 3.6 = 12.57\ cm^2$$

Coefficient of loss of power of the Secondary (P_s = 136 W), from the diagram: $1 + \frac{\Delta U}{U_s} = \upsilon_2 = 1.075$

*N_s – Number of turns of the Secondary_*

$$N_{s1} = \frac{U_{s1}(1 + \frac{\Delta U}{U_{s1}}) \times 10^4}{4.44 \times f \times B_m \times S_{Fe_{eff}}} = \frac{6.3 \times 1.075 \times 10^4}{4.44 \times 50 \times 1 \times 12.57} = 24.26$$

$$N_{s2} = \frac{U_{s2}(1 + \frac{\Delta U}{U_{s2}}) \times 10^4}{4.44 \times f \times B_m \times S_{Fe_{eff}}} = \frac{12.6 \times 1.075 \times 10^4}{4.44 \times 50 \times 1 \times 12.57} = 48.52$$

$$N_{s3} = \frac{U_{s3}(1 + \frac{\Delta U}{U_{s1}}) \times 10^4}{4.44 \times f \times B_m \times S_{Fe_{eff}}} = \frac{350 \times 1.075 \times 10^4}{4.44 \times 50 \times 1 \times 12.57} = 1348$$

N_p – Number of turns of the Primary

$$N_p = \frac{U_p \times 10^4}{4.44 \times f \times B_m \times S_{Fe}} = \frac{220 \times 10^4}{4.44 \times 50 \times 1 \times 12.57} = 788$$

Wire diameter (J = 2.5 A / mm²):

$d_{s1} = 1.13 \times \sqrt{\frac{I_{s1}}{J}} = 1.13 \times \sqrt{\frac{1.5}{2.5}} = 0.875\ mm;$ *standard Cul Ø 9 mm (insulated Cul Ø* 0.99 *mm)*

$d_{s2} = 1.13 \times \sqrt{\frac{I_{s2}}{J}} = 1.13 \times \sqrt{\frac{1}{2.5}} = 0.71\ mm;$ *standard Cul Ø 0.71 mm (insulated Cul Ø* 0.79 *mm)*

$d_{s3} = 1.13 \times \sqrt{\frac{I_{s3}}{J}} = 1.13 \times \sqrt{\frac{0.3}{2.5}} = 0.39\,mm;$ *standard Cul Ø 0.4 mm (insulated Cul Ø* 0.462 *mm)*

Wire diameter of the Primary winding (J = 2.5 A / mm^2*):*

$$I_p = \frac{P_p}{U_p \times \sqrt{\eta}} = \frac{130.2}{220 \times \sqrt{0.8}} = 0.66\ A$$

$d_p = 1.13 \times \sqrt{\frac{I_p}{J}} = 1.13 \times \sqrt{\frac{0.66}{2.5}} = 0.58\,mm$*; standard Cul Ø 0.6 mm (insulated Cul Ø* 0.675 *mm)*

Maximum length of winding:

$$\ell_c = c - (4 \times s) = 54 - 4 = 50\ mm$$

Number of turns per layer:

Maximum number of turns per layer, S$_1$*:*

$$N_{lc1} = 0.97 \times \frac{\ell_c}{d_{WP1}} = 0.97 \times \frac{50}{0.99} = 48.9\ turn/layer$$

One layer: 1 winding of 24 turns of wire Cul Ø 0.9 *mm and 1 winding of 24 turns of wire Cul Ø* 0.9 *mm.*
Standard Cul Ø 0.9 mm (insulated Cul Ø 0.99 *mm)*

Maximum number of turns per layer, S$_2$*:*

$$N_{lc2} = 0.97 \times \frac{\ell_c}{d_{WP2}} = 0.97 \times \frac{50}{0.79} = 61.3\ turn/layer$$

48 turns of S2 in one layer can be wound with insulated wire of maximum diameter layer:

$$d_{WP} = 0.97 \frac{\ell_c}{N_{S2}} = 0.97 \times \frac{50}{48.52} = 1\ mm$$

Standard Cul Ø 0.9mm (insulated Cul Ø 0.99 *mm)*

Maximum number of turns per layer, S$_3$*:*

$$N_{lc3} = 0.97 \times \frac{\ell_c}{d_{WP3}} = 0.97 \times \frac{50}{0.462} = 104\ turn/layer$$

Number of layers:

$$z_3 = \frac{N_p}{N_{lc3}} = \frac{1348}{104} = 12.96,\ \text{Integer = 13 layers}$$

S$_3$*: 13 layers x 104 turns = 1352 turns. Wire: Cul Ø 0.4 mm (insulated Cul Ø* 0.462 *mm)*

Maximum number of turns per layer, Primary P$_1$*:* $N_{lcp} = 0.97 \times \frac{\ell_c}{d_{WP}} = 0.97 \times \frac{50}{0.675} = 71$

Number of layers:

$$z_p = \frac{N_p}{N_{lcp}} = \frac{788}{71} = 10.9, \quad Integer = 11\ layers$$

P: 11 layers x 71 turns = 781 turns. Wire: Cul Ø 0.6 mm (insulated Cul Ø 0.675 mm)

Total radial dimension of the windings:

$$h_0 = 0.99 + 0.05 + 0.99 + 0.05 + 13 \times 0.462 + 0.3 + 11 \times 0.675 = 15.8\ mm$$

 # Condition: h_0(15.8 mm) < $h_c = a - s = (18 - 2) = 16$ mm <u>(condition is fulfilled)</u>

l_t *– mean length per turn (MLT):*

$$l_t = (2 \times b) + (2 \times h) + (\pi \times h_0) + (4 \times s) = (2 \times 3.6) + (2 \times 3.6) + (3.14 \times 1.58) + (4 \times 0.2) = 20.57\ cm$$

l_w *– wire length of the winding (m)*

$$l_{w_{s1}} = N_{s1} \times l_t = 24 \times 0.2057 = 4.93\ m$$

$$l_{w_{s2}} = N_{s2} \times l_t = 49 \times 0.2057 = 10\ m$$

$$l_{w_{s3}} = N_{s3} \times l_t = 1348 \times 0.2057 = 277.28\ m$$

$$l_{w_p} = N_p \times l_t = 788 \times 0.2057 = 162\ m$$

DC resistance of the windings:

$$R_{c_{s1}} = \rho \times \frac{4 \times l_{w_{s1}}}{\pi \times d^2} = 0.0175 \times \frac{4 \times 4.93}{3.14 \times 0.9^2} = 0.135\ \Omega$$

$$R_{c_{s2}} = \rho \times \frac{4 \times l_{w_{s2}}}{\pi \times d^2} = 0.0175 \times \frac{4 \times 10}{3.14 \times 0.9^2} = 0.275\ \Omega$$

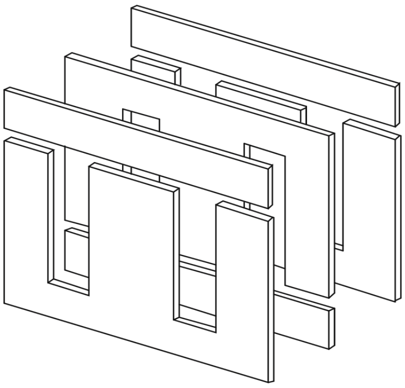
$$R_{c_{s3}} = \rho \times \frac{4 \times l_{w_{s3}}}{\pi \times d^2} = 0.0175 \times \frac{4 \times 277.28}{3.14 \times 0.4^2} = 38.6\ \Omega$$

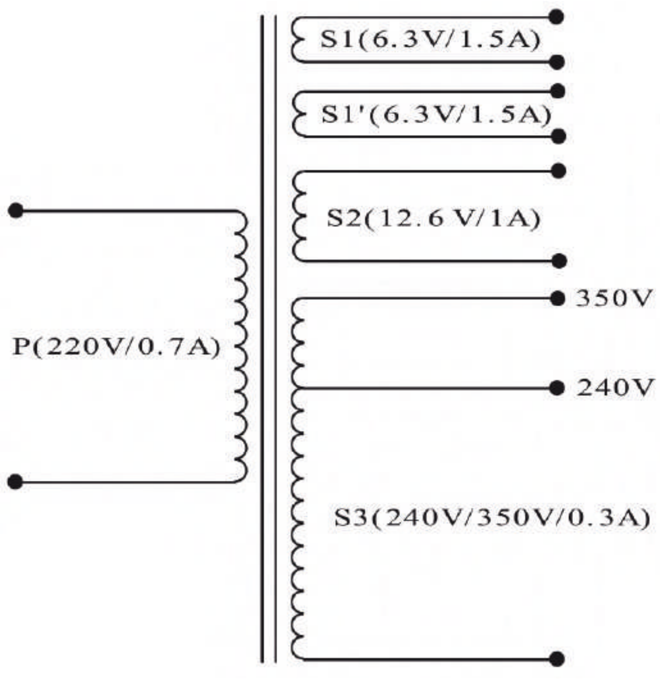
$$R_{c_p} = \rho \times \frac{4 \times l_{w_p}}{\pi \times d^2} = 0.0175 \times \frac{4 \times 162}{3.14 \times 0.6^2} = 10\ \Omega$$

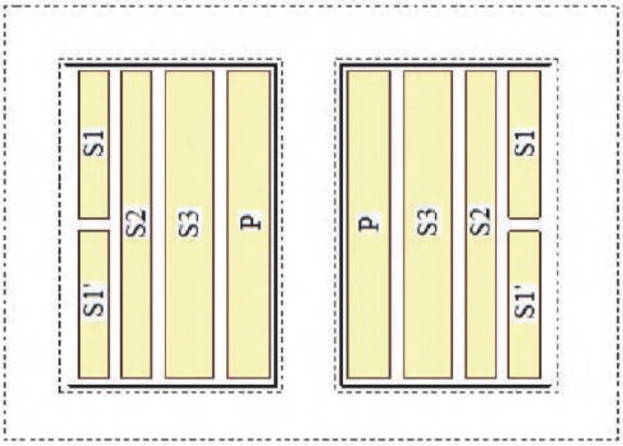
Winding procedure
Coil former (3.6 x 3.6) cm
EI 108 Fe 4 % Si NOSS transformer laminate Δ = 0.5 mm

1. *Primary winding: 11 layers of 71 turns each (total: 781 turns), wire Cul Ø 0.6 mm.*
2. *Insulation foil : # 0.3 mm*
3. *S3, Secondary winding: 9 layers of 104 turns each (total: 936 turns), wire Cul Ø 0.4 mm, per layer. TAP (240V) + 4 layers of 104 turns each (total: 416 turns), wire Cul Ø* 0.4 *mm.*
   S3 Secondary winding, total: 13 layers of 104 turns each (total: 1352 turns. Calculated number of turns: 1348. The difference is negligible. If the exact number of turns is desired, two of the 13 layers can be wound with 102 turns, for example the ninth and thirteenth layers), wire Cul Ø 0.4 mm.
4. *Insulation foil : # 0.0 5mm*
5. *S2, Secondary winding: 1 layer of 48 turns, Cul Ø 0.9 mm.*
6. *Insulation foil : # 0.0 5mm*
7. *S1, Secondary winding: 1 layer, 2 windings, 24 turns per winding, wire: Cul Ø 0.9 mm.*
8. *Insulation foil : # 0.0 5mm*

E laminates are stacked one by one in the opposite direction. ***I*** *laminates are inserted between* ***E*** *laminates.*







POWER SUPPLY TRANSFORMER FOR APPLICATION IN HIGH-END EQUIPMENT

For applications in Highi End apparatus, significantly higher quality power transformers than those used in consumer products are required, so much stricter values for maximum induction, current density and coefficient of weight ratio of iron core and cooper winding are used for the calculation:

- Induction $B_{max} = 0.9$ T
- Electric Current Density J = 2A / mm^2
- Coefficient of weight ratio of iron core and cooper winding G = 1.25
- Transformer efficiency $\eta = 0.8$

Cross-section of the transformer core (B_{max} = 9000 Gauss, J = 2A / mm^2, G = 1.25):

$$S_{Fe(min)}(cm^2) = \sqrt{\frac{P_{0_{maxac}} \times G \times 10^6}{\eta \times B \times J \times f}} = \sqrt{\frac{P_{0_{maxac}} \times 1.25 \times 10^6}{0.8 \times 9000 \times 2 \times f}} = 9.34 \times \sqrt{\frac{P_{0_{maxac}}}{f}}$$

Simplified equation:

$$S_{Fe(min)}(cm^2) = 9.34 \times \sqrt{\frac{P_{0_{maxac}}}{f}}$$

For f = 50 Hz:

$$S_{Fe(min)}(cm^2) = 9.34 \times \sqrt{\frac{P_{0_{maxac}}}{f}} = 9.34 \times \sqrt{\frac{P_{0_{maxac}}}{50}} = 1.32 \times \sqrt{P_{0_{maxac}}} \qquad (501.62)$$

Where the P_{0max} is the total power of all Secondaries:

$$P_{0max} = P_s = (U_{s1} \times I_{s1}) + (U_{s2} \times I_{s2}) + \cdots (U_{sn} \times I_{sn}) = \sum_{i=1}^{n} (U_{si} \times I_{si})$$

Core laminate:
Calculation of the optimal dimension of the transformer laminates (b):

$$b_{(max)}(cm) = \sqrt{S_{Fe}}$$

$$b_{(min)}(cm) = 0.85 \times \sqrt{S_{Fe}} = 0.85 \times b_{max}$$

$$b_{(min)} < b < b_{(max)}$$

Choose a standard transformer laminate

Calculation of the height of the stack of transformer laminates (h):

$$h(cm) = \frac{S_{Fe}(cm^2)}{k_i \times b(cm)}$$

Choose a standard coil former (or make it if it is not of standard dimensions).

k_i – stacking coefficient (0.92 – 0.98)
EI laminate Fe 4 % Si, $\Delta = 0.5$ mm
$k_i = 0.97$ (EI laminate $\Delta = 0.5$ mm)

Turn per Volt, simplified equation:

$$\frac{Turns}{V} = \frac{2250}{S_{Fe\,eff} \times f}$$

For f = 50 Hz:

$$\frac{Turns}{V} = \frac{45}{S_{Fe\,eff}}$$

N_p – Number of turns of the Primary:

$$N_P = U_P \times \frac{Turns}{V}$$

N_s – Number of turns of the Secondary:

$$N_s = U_s \times \frac{Turns}{V} \times \frac{1}{\sqrt{\eta}} = 1.118 \times U_s \times \frac{Turns}{V}$$

Wire diameter:

Wire diameter of the Primary winding (J = 2 A / mm^2):

$$I_p = \frac{P_p}{U_p \times \sqrt{\eta}} = 1.118 \times \frac{P_p}{U_p}$$

$$d_p = 1.13 \times \sqrt{\frac{I_p}{J}} = 1.13 \times \sqrt{\frac{I_p}{2A/mm^2}} = 0.8 \times \sqrt{I_P}$$

Wire diameter of the Secondary winding:

$$d_s = 0.8 \times \sqrt{I_S}$$

Further calculation is similar to the transformer calculation already explained (adjusting N_P and N_S as an integer, number of layers, turns per layer, standard diameter of Cu wire according to the possibility of fitting the windings in the window of the transformer core, DC resistance of the Primary and Secondary windings, Cu loss, etc).

Example:
 # Simplified and quick calculation

Mains voltage: 230V / 50Hz
Secondary:
S_1 *- 6.3 V / 4 A*
S_2 *- 6.3 V / 4 A*
S_3 *- 2 x 12.6 V / 1A*
S_4 *- 80 V / 0.06 A*
S_5 *- 360 V / 0.5 A (300 V + 10 V + 10 V + 10 V + 10 V + 10 V + 10 V)*

Total power of all Secondaries:

$$P_s = (U_{s1} \times I_{s1}) + (U_{s2} \times I_{s2}) + (U_{s3} \times I_{s3}) + (U_{s4} \times I_{s4}) + (U_{s5} \times I_{s5})$$

$$P_s = (6.3 \times 4) + (6.3 \times 4) + (2 \times 12.6 \times 1) + (80 \times 0.06) + (360 \times 0.5) = 260.45\ W$$

Cross-section of the transformer core, simplified equation:

$$S_{Fe(min)}(cm^2) = 9.34 \times \sqrt{\frac{P_{0_{maxac}}}{f}} = 9.34 \times \sqrt{\frac{260.45}{50}} = 21.31\ cm^2$$

Core laminate:
Calculation of the optimal dimension of the transformer laminates (b):

$$b_{(max)}(cm) = \sqrt{S_{Fe}} = \sqrt{21.31} = 4.6\ cm$$

$$b_{(min)}(cm) = 0.85 \times \sqrt{S_{Fe}} = 0.85 \times b_{max} = 0.85 \times 4.6 = 3.91\ cm$$

$$b_{(min)} < b < b_{(max)}$$

$$3.91 < b < 4.6$$

Standard transformer laminate: ***EI 120, b = 4 cm***

Height of stack of transformer laminates (h):

$$h(cm) = \frac{S_{Fe}(cm^2)}{k_i \times b(cm)} = \frac{21.31}{0.97 \times 4} = 5.49\ cm$$

EI laminate Fe 4 % Si, Δ = 0.5 mm.
k_i – 0.97 (EI laminate Δ = 0.5 mm)

Standard coil former: ***4 cm x 5.5 cm***

Effective cross-section of the transformer core:

$$S_{Fe\,eff} = k_i \times b \times h \times = 0.97 \times 4 \times 5.5 = 21.34\ cm^2$$

Turn per Volt:

$$\frac{Turns}{V} = \frac{2250}{S_{Fe\,eff} \times f}$$

For f = 50 Hz:

$$\frac{Turns}{V} = \frac{45}{S_{Fe\,eff}} = \frac{45}{21.34} = 2.1087$$

N_p – Number of turns of the Primary:

$$N_P = U_P \times \frac{Turns}{V} = 230 \times 2.1087 = 485\ turns$$

N_s – Number of turns of the Secondary:

$$N_s = U_s \times \frac{Turns}{V} \times \frac{1}{\sqrt{\eta}} = 1.118 \times U_s \times \frac{Turns}{V}$$

***N_{S1} = N_{S2}** :* *N_{S1} = N_{S2} = 1.118 x 6.3 x 2.1087 = 14.85 turns; INTEGER: 15 turns*

N_{S3} : *N_{S3} = 2 x (1.118 x 12.6 x 2.1087) = 2 x 29.7 turns; INTEGER: 2 x 30 turns*

N_{S4} : *N_{S4} = 1.118 x 80 x 2.1087 = 188.6 turns; INTEGER: 189 turns*

N_{S5} : *N_{S5} = 1.118 x 360 x 2.1087 = 848.7 turns; INTEGER: 849 turns*

Wire diameter:
Wire diameter of the Primary winding (J = 2 A / mm^2):

$$I_p = \frac{P_p}{U_p \times \sqrt{\eta}} = 1.118 \times \frac{P_p}{U_p} = 1.118 \times \frac{247.85}{230} = 1.0\ A$$

$$d_p = 1.13 \times \sqrt{\frac{I_p}{J}} = 1.13 \times \sqrt{\frac{I_p}{2A/mm^2}} = 0.8 \times \sqrt{I_P} = 0.8\sqrt{1} = 0.8\ mm$$

Standard ***Cul Ø 0.8 mm (****insulated* ***Cul*** Ø 0.885 mm)

Wire diameter of the Secondary winding:

$$d_s = 0.8 \times \sqrt{I_S}$$

S1, S2: $\mathbf{d_{S1}} = \mathbf{d_{S2}} = 0.8 \times \sqrt{4} = 1.6\ mm; Stabdard\ Cul\ Ø\ \mathbf{1.6\ mm}\ (insulated\ Cul\ Ø\ 1.711\ mm\ )$

S3: $\mathbf{d_{S3}} = 0.8 \times \sqrt{1} = 0.8\ mm; Stabdard\ Cul\ Ø\ \mathbf{0.8\ mm}\ (insulated\ Cul\ Ø\ 0.885\ mm\ )$

S4: $\mathbf{d_{S4}} = 0.8 \times \sqrt{0.06} = 0.196\ mm; Stabdard\ Cul\ Ø\ \mathbf{0.2\ mm}\ (insulated\ Cul\ Ø\ 0.245\ mm\ )$

S5: $\mathbf{d_{S5}} = 0.8 \times \sqrt{0.5} = 0.565\ mm; Stabdard\ Cul\ Ø\ \mathbf{0.56\ mm}\ (insulated\ Cul\ Ø\ 0.632\ mm\ )$

Maximum length of winding:

$$\ell_c = c - (4 \times s) = 60 - 4 = 56\ mm$$

Maximum number of turns per layer:

S_1: $N_{lc1} = 0.97 \times \frac{\ell_c}{d_{W\ S1}} = 0.97 \times \frac{56}{1.711} = 31.74\ turn/layer$

*Length of the winding, **S1**:* $\ell_{c_{S1}} = \frac{N_{S1} \times d_{W\ S1}}{k_i} = \frac{15 \times 1.711}{0.97} = 26.46\ mm$

*Length of the winding, **S2**:* $\ell_{c_{S2}} = \frac{N_{S2} \times d_{W\ S2}}{k_i} = \frac{15 \times 1.711}{0.97} = 26.46\ mm$

$$\ell_{c_{S1}} + \ell_{c_{S2}} = 26.46 + 26.46 = 52.92\ mm < \ell_c = 56\ mm$$

As can be seen from the above calculation, the windings of both secondary S1 and S2 can be wound in one layer.

***S3**:* $\ell_{c_{S3}} = 2 \times \left(\frac{N_{S3} \times d_{W\ S3}}{k_i}\right) = 2 \times \frac{30 \times 0.885}{0.97} = 2 \times 27.37 = 54.7\ mm$

***S4**:* $\ell_{c_{S4}} = \left(\frac{N_{S4} \times d_{W\ S4}}{k_i}\right) = \frac{189 \times 0.245}{0.97} = 47.7\ mm$

In order for the winding to completely fill one layer, use the standard Cul Ø 0.236 *mm (insulated Cul Ø* 0.285 *mm,) wire.*

***S5**:*
Turns per layer:

$$N_{lc5} = 0.97 \times \frac{\ell_c}{d_{W\ S5}} = 0.97 \times \frac{56}{0.632} = 85.9 \frac{turn}{layer}; INTEGER\ \ 85 \frac{turn}{layer}$$

Number of layers:

$$z_5 = \frac{N_{S5}}{N_{lc5}} = \frac{849}{85} = 9.98;$$ *INTEGER = 10 layers*

S5: 10 layers of 85 turns each (total: 850 turns), wire Cul Ø 0.56 mm, (TAPS: 708 turns + 24 turns +24 turns + 24 turns + 24 turns + 23 turns + 23 turns).

Primary, P:

Turns per layer:

$$N_{lcP} = 0.97 \times \frac{\ell_c}{d_{W\ P}} = 0.97 \times \frac{56}{0.885} = 61.3 \frac{turn}{layer}; INTEGER\ \ 61 \frac{turn}{layer}$$

Number of layers:

$$z_P = \frac{N_P}{N_{lcP}} = \frac{485}{61} = 7.95;\ \ INTEGER = 8\ layers$$

P: 8 layers of 61 turns (total: 488 turns), wire Cul Ø 0.8 mm.

Total radial dimension of the windings:

$$\boldsymbol{h_0} = 1.711 + 0.05 + 0.885 + 0.05 + 0.285 + 0.05 + 10 \times 0.632 + 0.3 + 8 \times 0.885 + 16 \times 0.05 = \mathbf{17.53}\ \boldsymbol{mm}$$

*Condition:* $h_0(17.53\ mm) < h_c = a - s = (20 - 2) = 18\ mm$ *(condition is fulfilled)*

l_t **–** *mean length per turn (MLT):*

$$\boldsymbol{l_t = (2 \times b) + (2 \times h) + (\pi \times h_0) + (4 \times s) = (2 \times 4) + (2 \times 5.5) + (3.14 \times 1.673) + (4 \times 0.2) = 25\ cm}$$

l_w **–** *wire length of the winding (m):*

$l_W = N \times l_t$

$$l_{w_{s1}} = N_{s1} \times l_t = 15 \times 0.25 = 3.75\ m$$
$$l_{w_{s2}} = N_{s2} \times l_t = 15 \times 0.25 = 3.75\ m$$
$$l_{w_{s3}} = N_{s3} \times l_t = 30 \times 0.25 = 7.5\ m$$
$$l_{w_{s4}} = N_{s4} \times l_t = 189 \times 0.25 = 47.25\ m$$
$$l_{w_{s5}} = N_{s5} \times l_t = 850 \times 0.25 = 212.5\ m$$
$$l_{w_p} = N_p \times l_t = 488 \times 0.25 = 122\ m$$

DC resistance of the windings:

$$R_{c_{s1}} = R_{c_{s2}} = \rho \times \frac{4 \times l_{w_{s1}}}{\pi \times d_{S1}^2} = 0.0175 \times \frac{4 \times 3.75}{3.14 \times 1.6^2} = 0.0326\ \Omega$$

$$R_{c_{s3}} = \rho \times \frac{4 \times l_{w_{s3}}}{\pi \times d_{S3}^2} = 0.0175 \times \frac{4 \times 7.5}{3.14 \times 0.8^2} = 0.26\ \Omega$$

$$R_{c_{s4}} = \rho \times \frac{4 \times l_{w_{s4}}}{\pi \times d_{S41}^2} = 0.0175 \times \frac{4 \times 47.25}{3.14 \times 0.2^2} = 26.32\ \Omega$$

$$R_{c_{S5}} = \rho \times \frac{4 \times l_{w_{S5}}}{\pi \times d_{S5}^2} = 0.0175 \times \frac{4 \times 212.5}{3.14 \times 0.56^2} = 15\ \Omega$$

$$R_P = \rho \times \frac{4 \times l_{wP}}{\pi \times d_P^2} = 0.0175 \times \frac{4 \times 122}{3.14 \times 0.8^2} = 4.247\ \Omega$$

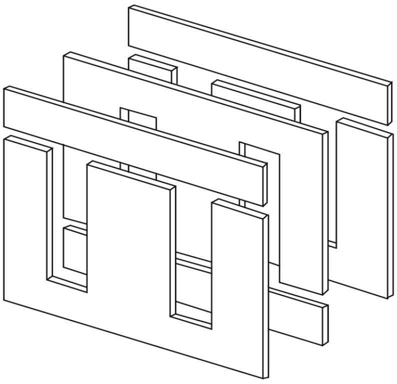
Winding procedure

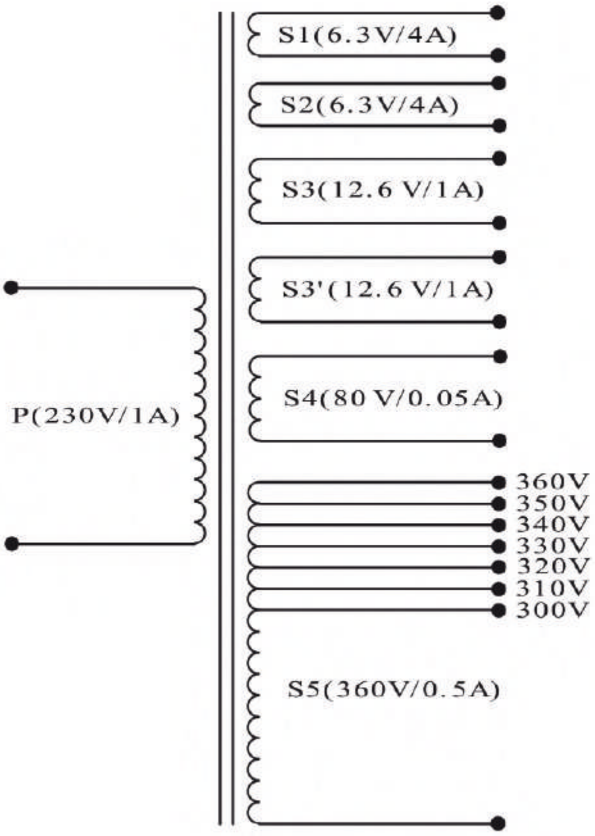
Coil former (4 x 5.5) cm

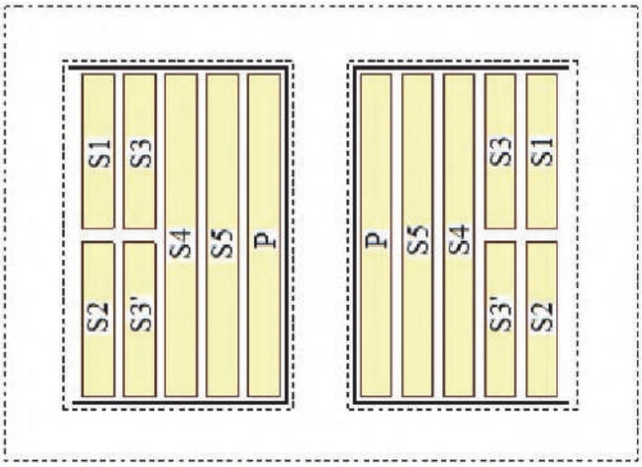
EI 120 Fe 4 % Si NOSS transformer laminate Δ = 0.5 mm

1. *Primary, P: 8 layers of 61 turns each (total: 488 turns), wire Cul Ø 0.8 mm, insulation foil between layers: 0.05mm.*
2. *Insulation foil: # 0.3 mm*
3. *Secondary, S5: 10 layers of 85 turns each (total: 850 turns), wire Cul Ø 0.56 mm, (TAPS: as explained above), insulation foil between layers: # 0.05 mm.*
4. *Insulation foil: # 0.05 mm*
5. *Secondary, S4: 1 layer of 189 turns, wire Cul Ø 0.236 mm.*
6. *Insulation foil: # 0.05 mm*
7. *Secondary, S3: 1 layer, 2 windings of 30 turns each, wire Cul Ø 0.8mm.*
8. *Insulation foil: # 0.05 mm*
9. *Secondary, S1, S2: 1 layer 2 windings of 15 turns each, wire Cul Ø 1.6 mm*
10. *Insulation foil: # 0.05 mm*

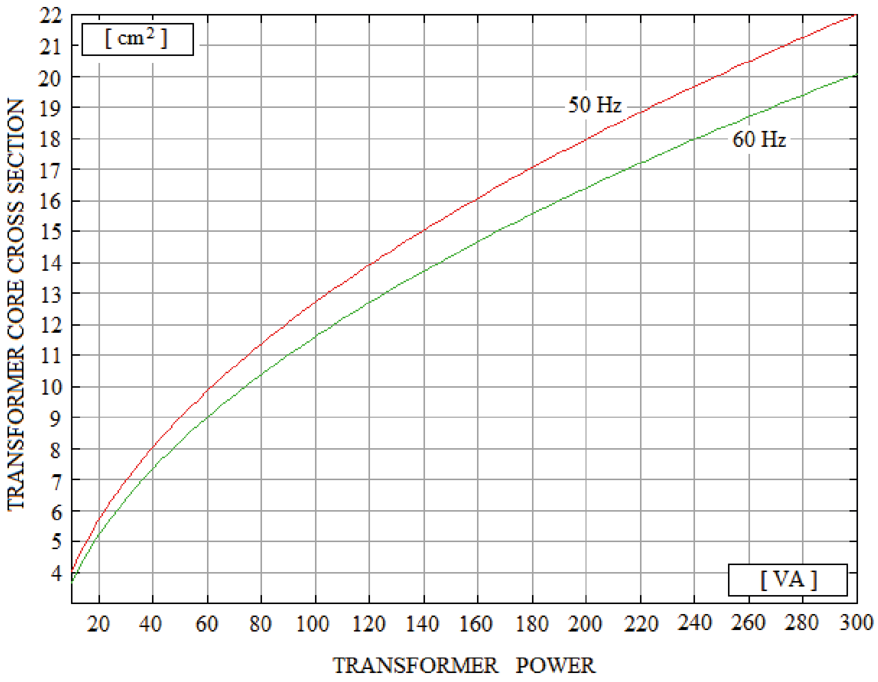
E laminates are stacked one by one in the opposite direction. ***I*** *laminates are inserted between* ***E*** *laminates.*




S1(6.3V/4A)
S2(6.3V/4A)
S3(12.6 V/1A)
S3'(12.6 V/1A)
S4(80 V/0.05A)
P(230V/1A)
360V
350V
340V
330V
320V
310V
300V
S5(360V/0.5A)


S1
S3
S4
S5
P
S2
S3'
P
S5
S4
S3
S1
S3'
S2

A useful diagram for quickly determining the power transformer core cross section



Thermal characteristics of the power transformer

A part of the transformer energy (in coils and iron core) is converted into thermal energy and heats the transformer. The thermal energy of the transformer is transferred to the environment through the outer surfaces of the transformer. After a certain operation of the transformer, thermal equilibrium is established. In conditions of thermal equilibrium, the temperature of the transformer is constant. When calculating and designing a transformer, it is very important to predict the temperature of the transformer in operating conditions. Since the operating temperature of the transformer depends on the ambient temperature, the calculation of the transformer temperature increase in relation to the ambient temperature is used in the transformer calculation. The calculation is based on the power loss of the transformer (power loss in the iron core and transformer windings).

Iron core heat loss:

The heat loss of the transformer core depends on the type and mass of material used.

1. Heat loss of the iron core per mass unit **W / kg**:

   NOSS E/I, 0.35 mm: 1.07 ... 1.21 W / kg (1T, 50 Hz)
   NOSS E/I, 0.50 mm: 1.20 ... 1.28 W / kg (1T, 50 Hz)
   GOSS E/I, 0.50 mm: 0.42 W / kg (1T, 50 Hz)

2. The mass of the iron core:

   $$G_{Fe}(gr) = V_{Fe}(cm^3) \times D_{Fe}\left(\frac{g}{cm^3}\right) = S_{eff}(cm^2) \times \ell_{Fe}(cm) \times D_{Fe}\left(\frac{g}{cm^3}\right)$$

   D_{Fe} – ferromagnetic material density

   (Most commonly used material: **4 % Si Ferro.** Density: **$D_{Fe(4\%Si)}$ = 7.65 g / cm³).**

   l_{Fe} - Magnetic circuit (path) length MPL $\ell_{Fe}(cm) = 5.57 \times b(cm)$

   S_{Fe} – cross-section of the transformer core

3. Iron core heat loss

   $$P_{Fe}(W) = G_{Fe}(kg) \times \frac{W}{kg}\ [W]$$

Heat losses of transformer windings, Cu loss in Primary and Secondary windings:

1. DC resistance of the Primary winding:

   $R_{P\,DC} = \rho \times \frac{4 \times l_w}{\pi \times d^2}$; d – wire diameter (mm), ρ = specific resistance of the wire = 0.0175 $\Omega mm^2/m$,

   l_w – length of the winding wire (m): l_w = N x l_t **; N –** number of turns**.**

   l_t**–** mean length per turn (ℓ_t):

   $$l_t = (2 \times b) + (2 \times h) + (\pi \times h_0) + (4 \times s)$$

2. Power loss of the Primary winding:

   $P_{P-Cu\,DC}(W) = R_{P\,DC} \times I_P^2\ [W]$ **;** I_P – primary current

3. Power losses of the Secondary windings:

   $\sum_{i=1}^{n} P_{S(i)-\,Cu\ DC} = \sum_{i=1}^{n} R_{S(i)\,DC} \times I_{S(i)}^2$; Total power loss of all secondary windings.

Total power loss of the transformer (total heat power):

The sum of the power losses of the iron core and the primary and all the secondary windings

$$P_{loss\ total} = P_{Fe} + P_{P-Cu\,dc} + \sum_{i=1}^{n} P_{S(i)-\,Cu\ dc}$$

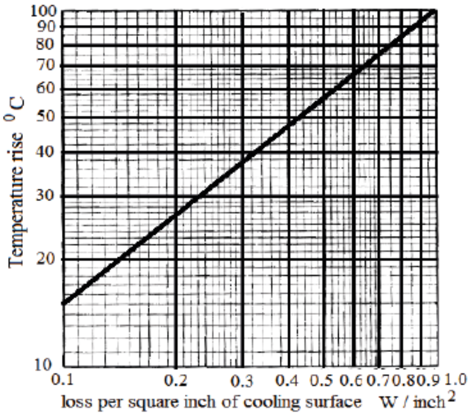
Heat radiation surface of the transformer – cooling surface (E / I transformer laminate):
* Equation published in Radiotron Designer's Handbook Fourth Edition in 1953. (RDH4):

$$S_{heat\ rad} = b \times [(7.71 \times b) + (11 \times h)]$$

** Mr. Patrick Turner, in his excellent articles (Turner Audio web site), published a slightly modified form of the above equation. Mr. Patrick's opinion is that the entire outer surface of the transformer radiates thermal energy (which makes sense) and not just the outer surface of the iron core as published in RDH4.

$$S_{heat\ rad\ Patrick} = b \times [(15 \times b) + (11 \times h)]$$

*** The author of this book has no reason not to trust Mr. Patrick's equation, especially because of the results of measurements on transformers made and published by Mr. Patrick, but, for the sake of methodological correctness, the author reserves a degree of reservations regarding the application of Mr. Patrick's equation to the diagram: Temperature rise versus loss per square inch of cooling surface for ambient temperature of 25^0C. (The diagram is probably empirical and is valid for the cooling surface area calculated as in the original equation *).



Calculation:

$P_{loss\ total}\ [W]\ /\ S_{heat\ rad}[inch^2]$;

$S_{heat\ rad}$ must be calculated in $inch^2$, or if it is calculated in cm^2 it must be converted to $inch^2$.

Using the diagram:
Temperature rise versus loss per square inch of cooling surface of power transformer for ambient temperature of 25^0.

Note:
The heat radiation (cooling) surface of the power transformer must be expressed in square inches**.**

Reading the temperature rise (T_{rise}) from the diagram for the calculated:

$$P_{loss\ total}\ [W]\ /\ S_{heat\ rad}[inch^2]$$

Expected operating temperature of the power transformer:
The operating temperature of the power transformer is higher than the ambient temperature for the T_{rise} read from the diagram above:

$$T_{operation}[°C] = T_{ambient}[°C] + T_{rise}[°C]$$

Example:
Power transformer (previous example, P = 260W)
Mains voltage: 230V / 50Hz
Secondary:
S_1 - 6.3 V / 4 A
S_2 - 6.3 V / 4 A
S_3 - 2 x 12.6 V / 1 A
S_4 - 80 V / 0.06 A
S_5 - 360 V / 0.5 A (300 V + 10 V + 10 V + 10 V + 10 V + 10 V + 10 V)
E I 120, Fe 4 % Si NOSS transformer laminate Δ = 0.5 mm.
Coil former 4 cm x 5.5 cm.
Cross-section of the transformer core: $S_{eff} = 21.34\ cm^2$.
DC resistance of the windings: $R_{P-DC} = 4.247\ Ω$, $R_{S1-DC} = 0.0326\ Ω$, $R_{S2-DC} = 0.0326\ Ω$, $R_{S3-DC} = 0.26\ Ω$, $R_{S4-DC} = 26.32\ Ω$, $R_{S5-DC} = 15\ Ω$.

Cu loss in windings:

$P_{loss} = R_{DC} \times I^2$

$P_{S1\ loss} = 0.032 \times 4^2 = 0.512\ W$; $P_{S2\ loss} = 0.032 \times 4^2 = 0.512\ W$; $P_{S3\ loss} = 0.26 \times 1^2 = 0.26\ W$;
$P_{S4\ loss} = 26.32 \times 0.06^2 = 0.094\ W$; $P_{S5\ loss} = 15 \times 0.5^2 = 3.75\ W$; $P_{P\ loss} = 4.247 \times 1^2 = 4.247\ W$;
$P_{Cu\ loss\ total} = 9.375\ W$;

Mass of iron core:

$$G_{Fe}(gr) = V_{Fe}(cm^3) \times D_{Fe}\left(\frac{g}{cm^3}\right) = S_{eff}(cm^2) \times \ell_{Fe}(cm) \times D_{Fe}\left(\frac{g}{cm^3}\right)$$

$$G_{Fe}(gr) = S_{eff}(cm^2) \times l_{Fe}(cm) \times D_{Fe}\left(\frac{g}{cm^3}\right) = 21.34 \times (5.57 \times 4) \times 7.56 = 3.637\,kg$$

Iron core heat loss:

$P_{Fe\ loss}$ = G_{Fe} *x Heat loss of the iron core per mass unit (W/ kg) = 3.637 kg x 1.2 W/ kg (*NOSS E/I, 0.50 mm) = ***4.36 W***

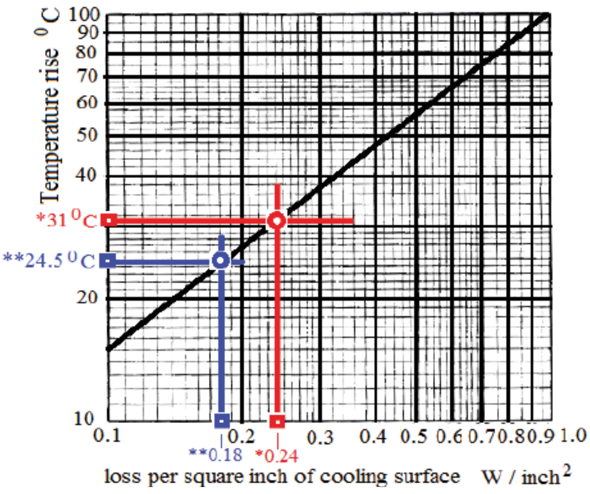
Total loss of the transformer:

$P_{loss\ total}$ = $P_{Cu\ loss\ total} + P_{Fe\ loss}$ = 9.375 + 4.36 =***13. 74W***

Heat radiation surface of the transformer (E / I transformer laminate):

** **$S_{heat\ rad}$** = b x [(11xh) + (7.71xb)] = 4 x [(11x5.5) + (7.71x4)] =365.36 cm^2 =* ***56.63 $inch^2$****;*

*** **$S_{heat\ rad\ Patrick}$** = b x [(11xh) + (15xb)] = 4 x [(11x5.5) + (15x4)] =482 cm^2 =* ***74.7 $inch^2$****;*



*** $P_{loss\ total} / S_{heat\ rad}$ = 13.74 / 56.63 = 0.24 W / $inch^2$;***

**** $P_{loss\ total} / S_{heat\ rad\ Patric}$ = 13.74 / 74.7 = 0.18 W / $inch^2$;***

$T_{amb} = 25^0C$

Temperature rise (T_{rise}) read from the diagram:

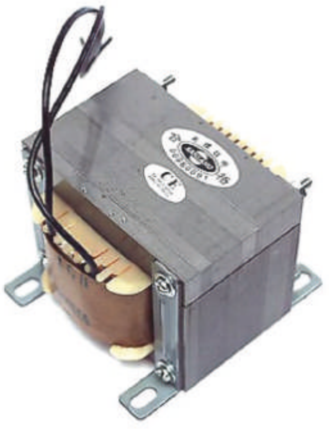
*** $T_{rise} = 31^0C$; ** $T_{rise} = 24.5^0C$***

* ***$T_{operation}$*** = $T_{amb} + T_{rise}$ = 25 + 31 = ***56 ^{0}C***

** ***$T_{operation}$*** = $T_{amb} + T_{rise}$ = 25 + 24.5 = ***49.5 ^{0}C***

The expected operating temperature of the transformer is 56 ^{0}C (49.5 ^{0}C).

5.8 CHOKES



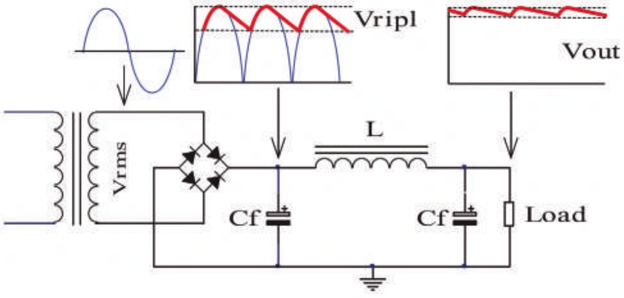
A choke is a passive electronic component. Generally, it is inductance. In practice, it is a winding made of Cu wire wound on an air-gapped iron core of the transformer. Like any inductance, a choke opposes the flow of alternating current in an electric circuit.

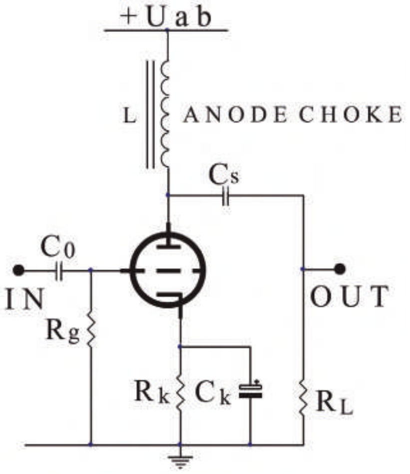
In tube audio devices, chokes are most commonly used in power supply circuits as filter components that reduce noise and hum (ripple), and have no significant effects on the DC current flowing through the circuit. Also, any sudden current changes occur in the circuit are smoothed out by the energy stored in the choke in the form of a magnetic field and, of course, by the electrical energy stored in the filter capacitors.

It can be used as a plate (anode) load of the tube, too.

Induction in the choke iron core is mainly caused by DC current (B_{DC}), but induction caused by the AC current (B_{AC}) cannot be ignored.

A good quality choke must be designed to have **low DC resistance** and **high inductance** as well as to **handle** the **required DC current**. **The induction** in the iron core must be low (in a **linear** part of the **B-H curve**).





All basic electromagnetic equations can be applied in the design and calculation of chokes.

Basic Equations

$$B_{DC} = \frac{1.257 \times N \times I \times \mu_e}{l_{Fe}}$$

$$B_{AC} = \frac{U_{RMS} \times 10^8}{4.44 \times N \times f \times S_{Fe}}$$

$$Induction\ in\ a\ choke\ iron\ core:\ B = (0.7\,T - 0.9\,T)$$

B_{DC} (Gauss) [Induction caused by the DC current]
B_{AC} (Gauss) [Induction caused by the AC current]
I (A) [Electric Current]
$l_{Fe}(\mathrm{cm})$; $(l_{Fe} = 5.57 \times b)$; [Magnetic circuit (path) length MPL]
$S_{Fe}(cm^2)$ [cross-section of the iron core]
$l_g(\mathrm{cm})$; Total air gap
$\delta(\mathrm{cm})$; Air gap
f(Hz) [Frequency]
J (A / mm^2) – electric current density (**J = 2.5 A / mm^2** is **recommended** for choke design)

Choke Inductance:

$$L = \frac{1.257 \times N^2 \times S_{Fe} \times \mu_{eff}}{l_{Fe} \times 10^8} \qquad (501.62)$$

$$\mu_{eff} = \frac{\mu_{max}}{1 + \mu_{max} \times \frac{l_g}{l_{Fe}}}; \qquad \delta = \frac{l_g}{2}; \qquad \left(l_g = \frac{l_{Fe} \times (\mu_{max} - \mu_e)}{\mu_{max} \times \mu_e}\right)$$

$\mu_{max} = 3000$, NOSS 4 % Si Fe laminate; $\mu_{eff} \leq 450$

DC resistance of the choke:

$$R_w = \rho \times \frac{4 \times l_w}{\pi \times d^2} = \rho \times \frac{4 \times N \times l_t}{\pi \times d^2}$$

d – wire diameter (mm)

ρ – specific resistance of the wire **(Cu wire: $\rho = 0.0175\,\Omega \frac{mm^2}{m}$)**
l_w **–** length of the winding wire (m): $l_w = N \times l_t$
N – number of turns**.**
l_t **–** mean length per turn (ℓ_t): $l_t = (2 \times b) + (2 \times h) + (\pi \times h_0) + (4 \times s)$

CHOKE DESIGN

A procedure of calculating choke using Hanna's diagram

Caution: The original Hanna's diagram uses **Imperial** units (inches).

This method is applied to design chokes with low AC flux density, which are used, for example, in CLC power supplly filters.

1. **Input data:** I_{DC} (A), L (H), f (Hz)

2. **Calculating wire diameter:**

$d_{min} = 1.13 \times \sqrt{\frac{I}{2.5}}$; Choose a standard wire diameter closest to d_{min}.

3. **The calculation starts with some standard Iron Core**

   Calculation: l_{FE}(inch) and S_{Fe} (square inch)

   Calculation: $\frac{L \times I^2}{V} = \frac{L \times I^2}{l_{Fe} \times S_{Fe}}$

4. **Reading** from **Hanna's diagram:** $\frac{N \times I}{l_{Fe}}$ and l_g / l_{Fe}

   Calculating the number of turns N using the **reading** $\frac{N \times I}{l_{Fe}}$ **:**

$$N = reading \left(\frac{N \times I}{l_{Fe}}\right) \times \frac{l_{Fe}}{I}$$

Calculating the total air gap l_g and δ using **reading** l_g / l_{Fe}**:**

$$l_g = reading \left(\frac{l_g}{l_{Fe}}\right) \times l_{Fe}$$

Calculating μ_{eff} : $\mu_{eff} = \frac{\mu}{1 + \mu \times \frac{\ell_g}{\ell_{Fe}}}$

5. **Checking the possibility of fitting the winding in the iron core window**

   Simplified calculation: $N\ calculated\ in\ 4. \leq \frac{0.64 \times S_W}{O_d^2}$

6. **Calculating the inductance L:** $L = \frac{1.257 \times N^2 \times S_{Fe} \times \mu_{eff}}{l_{Fe} \times 10^8}$

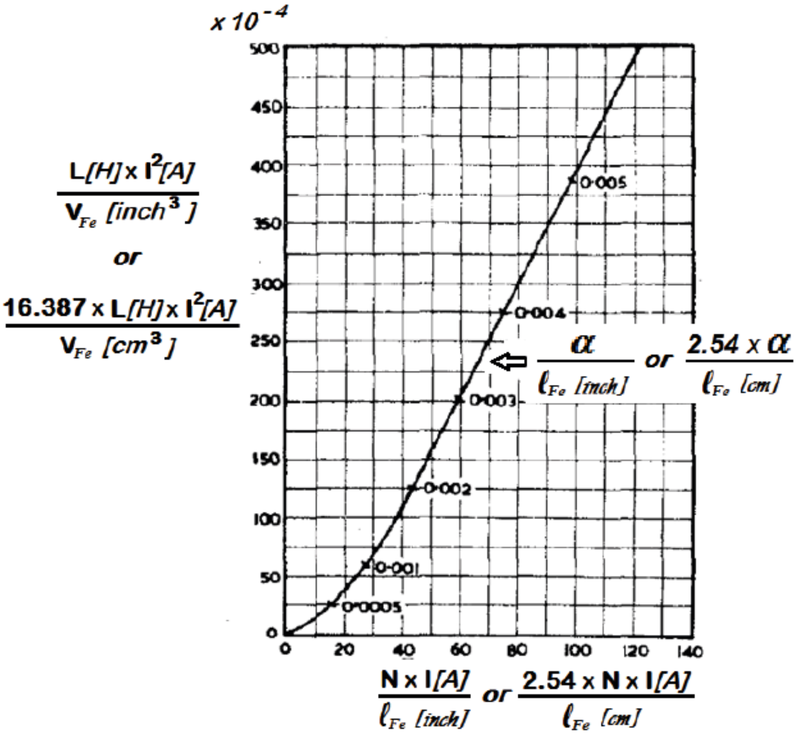
7. **Checking the DC flux density:** $B_{DC} = \frac{1.257 \times N \times I \times \mu_{eff}}{l_{Fe}}$ $\quad B_{DC} \leq (0.7\,T - 0.9\,T)$

8. **Calculating the DC resistance of the choke R_W using the equation:**

$$R_W = \rho \times \frac{4 \times N \times l_t}{\pi \times d^2} = \rho \times \frac{4 \times N \times (2 \times b) + (2 \times h) + (\pi \times h_0) + (4 \times s)}{\pi \times d^2}$$

9. **Calculating choke power loss:** $P_{loss} = R_W \times I_{DC}^2$

* **If the calculated L doesnt fullfill the required value or the induction in the iron core is too high, the calculation must be repeated using a standard Iron Core of a different size.**



Numerical method

1. **Input data:** I_{DC} (A), L (H), f (Hz)

2. **Calculating wire diameter:**

   $d_{min} = 1.13 \times \sqrt{\frac{I}{J}} = 1.13 \times \sqrt{\frac{I}{2.5}}$; Choose a standard wire diameter closest to d_{min}.

3. **The calculation starts with some standard Iron Core**

   Calculation: l_{Fe} (cm) and S_{Fe} (cm^2)

   Calculation: N [maximum number of turns that van be fitt in the coil former window], (turn / layer, number of layer or using the simplified equation**:** $N_{max} = \frac{0.64 \times S_W}{o_d^2}$).

   S_W [mm²] – cross-section of the iron core window
   O_d [mm] – diameter of insulated wire

   Calculating μ_{eff} using the equation B_{DC}= $\frac{1.257 \times N \times I \times \mu_{eff}}{l_{Fe}}$ **:** $\mu_{eff} = \frac{B_{DC} \times \ell_{Fe}}{1.257 \times N \times I}$

   B_{DC} = (7000 ÷ 9000) Gauss, μ_{eff} has to be $\leq$ 400 (450) .

4. **Calculating using the equation:** $l_g = l_{Fe} \times \frac{\mu_{max} - \mu_{eff}}{\mu_{max} \times \mu_{eff}}$ and $\delta = l_g / 2$

5. **Calculating the inductance L using the equation:** $L = \frac{1.257 \times N^2 \times S_{Fe} \times \mu_{eff}}{l_{Fe} \times 10^8}$

6. **Calculating R_W using the equation:** $R_W = \rho \times \frac{4 \times N \times l_t}{\pi \times d^2} = \rho \times \frac{4 \times N \times (2 \times b) + (2 \times h) + (\pi \times h_0) + (4 \times s)}{\pi \times d^2}$
   ρ = 0.0175 Ωmm²/*m*

7. Calculating choke power loss: $P_{loss} = R_W \times I_{DC}^2$

* If the calculated L doesnt fullfill the required value or the induction in the iron core is too high, the calculation must be repeated using a standard Iron Core of a different size.

Example:
CLC Choke
5H / 250 mA

Hanna's method

1. Input: *I_{DC} = 250 mA, L = 5 H, f = 50 Hz*

2. Calculating wire diameter:

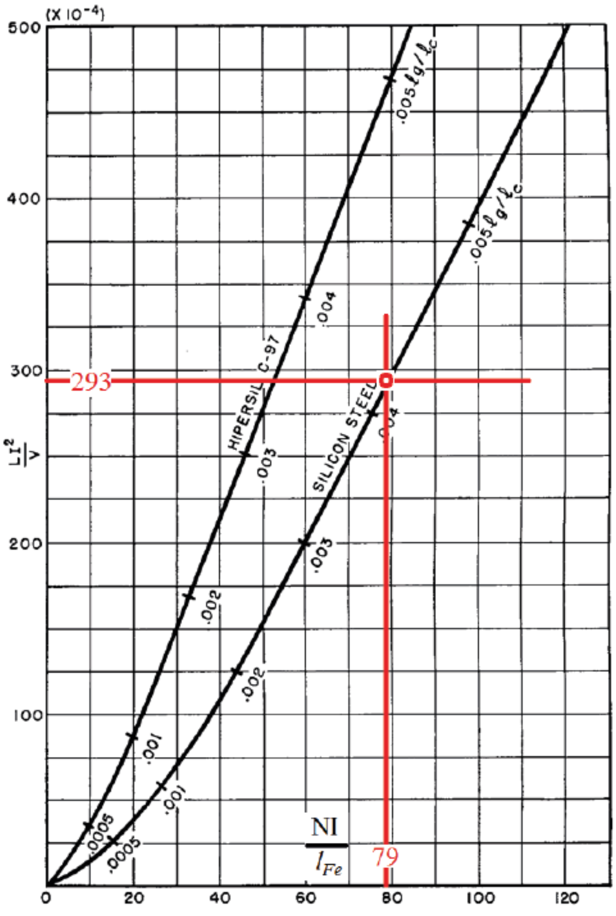
$$d_{min} = 1.13 \times \sqrt{\frac{I}{2.5}} = 1.13 \times \sqrt{\frac{0.25}{2.5}} = \mathbf{0.357\ mm}$$

Standard wire: ***Cul Ø 0.4 mm*** (***insulated:*** Ø 0.462 mm **=** O_d *)*

3. *The calculation starts with standard Iron Core* ***EI 96*** *NOSS Fe 4% Si Δ = 0.35 mm and* ***coil former (3.2 x 3.2) cm; [1.26 x 1.26] inch.***

ℓ_{Fe} (inch) = 5.57 x 1.26 = 7 inch = 17.824 cm;
S_{Fe} (square inch) = 0.96 x (1.26 x 1.26) = 1.524 inch^2 = 9.83 cm^2

$$\frac{L\times I^2}{V} = \frac{L\times I^2}{l_{Fe}\times S_{Fe}} = \frac{5\times 0.25^2}{7\times 1.524} = 293 \times 10^{-4}$$



4. Readings from the Hanna's diagram for $\frac{L\times I^2}{V} = 293 \times 10^{-4}$**:**

$$\frac{N\times I}{l_{Fe}} = 79 \quad and \quad \frac{l_g}{l_{Fe}} = 0.00415$$

Number of turns N, using the **reading** $\frac{N\times I}{l_{Fe}}$**:**

$$N = reading\left(\frac{N\times I}{l_{Fe}}\right) \times \frac{l_{Fe}}{I} = 79 \times \frac{7}{0.25} =$$

$$= \mathbf{2212}\ Turn$$

Total air gap l_g and δ, using the **reading** l_g **/** l_{Fe}**:**

$$l_g = reading\left(\frac{l_g}{l_{Fe}}\right) \times l_{Fe}$$

$$= 0.00415 \times 7 = \mathbf{0.029}\ inch = \mathbf{0.737\ mm}$$

$$\delta = l_g / 2 = 0.029 / 2 = \mathbf{0.0145\ inch} = \mathbf{0.369\ mm}$$

$$\mu_{eff}: \quad \mu_{eff} = \frac{\mu}{1+\mu\times\frac{l_g}{l_{Fe}}} = \frac{3000}{1+3000\times\frac{0.029}{7}} = \mathbf{223}\ ;$$

$$\mu_{eff} = 223 \leq 450$$

5. Checking the possibility of fitting the winding in the iron core window

Simplified calculation: $N\ calculated\ in\ 4.\ \leq \frac{0.64\times S_W}{O_d^2} = 2212 \leq \frac{0.64\times 768}{0.462^2} = 2302$

6. *Inductance L (using the classical equation):

$$L = \frac{1.257\times N^2\times S_{Fe}\times \mu_{eff}}{l_{Fe}\times 10^8} = \frac{1.257\times 2212^2\times 0.96\times 3.2\times 3.2\times 223}{5.57\times 3.2\times 10^8} = \mathbf{7.5\ H}$$

7. *Checking the DC flux density (using the classical equation):

$$B_{DC} = \frac{1.257\times N\times I\times \mu_e}{l_{Fe}} = \frac{1.257\times 2212\times 0.25\times 223}{5.57\times 3.2} = \mathbf{8697\ Gauss}$$

$$B_{DC} \leq (0.7T \div 0.9T)$$

8. Calculating the DC resistance of the choke R_W

$$R_W = \rho\times\frac{4\times N\times l_t}{\pi\times d^2} = \rho\times\frac{4\times N\times[(2\times b)+(2\times h)+(\pi\times h_0)+(4\times s)]}{\pi\times d^2} = 0.0175\times\frac{4\times N\times[2\times 0.032+2\times 0.032+3.14\times 0.013+4\times 0.002]}{3.14\times 0.4^2} = \mathbf{54.5\ \Omega}$$

9. Calculating choke power loss: $P_{loss} = R_W\times I_{DC}^2 = 54.5\times 0.25^2 = \mathbf{3.4\ W}$

Numerical method

1. Input data: I_{DC} = 250 mA, L = 5 H

2. Wire diameter:

$$d_{min} = 1.13\times\sqrt{\frac{I}{J}} = 1.13\times\sqrt{\frac{I}{2.5}} = 1.13\times\sqrt{\frac{0.25}{2.5}} = 0.357\ mm$$

Standard wire: Cul Ø 0.4 mm (Insulated: Ø 0.462 mm)

3. *The calculation starts with standard Iron Core* ***EI 96*** *NOSS Fe 4% Si Δ = 0.35 mm and* ***coil former (3.2 x 3.2) cm***

S_{Fe} *= 0.96 x 3.2 x 3.2 =* ***9.535 cm²***

l_{Fe} *= 5.57 x 3.2 =* ***17.824 cm***

V_{Fe} *= S_{Fe} x l_{Fe} = 9.535 x 17.824 = 169.96 cm³ (Imperial: 169.96 cm³ / 2.54³ = 10.37 cubic inch)*

Number of turns:

c” = (3 x a) – 4 mm = (3 x 16) – 4 = 44 mm
(n / layer) = 0.97 x (c” / O_d) = 0.97 x (44 / 0.462) = 92.38 Integer: 92 turns/layer
n_{layer} = h_0 /O_d = (16 - 4) / 0.462 = 25.97 Integer: 25 layers

N *= n_{layer} x (n / layer) = 92 x 25 =* ***2300 turns***

Simplified calculation: $\mathbf{N_{max}} = \frac{0.64\times S_W}{O_d^2} = \frac{0.64\times(16\times 48)}{0.462^2} = \mathbf{2303\ turns}$

Calculating μ_e for $B_{DC\ min}$ = *7000 Gauss:* $\mu_e = \frac{l_{Fe}\times B_{DC}}{1.257\times N\times I} = \frac{17.824\times 7000}{1.257\times 2300\times 0.25} = \mathbf{172.6}$

4. Air gap: $l_g = l_{Fe}\times\frac{\mu_{max}-\mu_e}{\mu_{max}\times\mu_e} = 17.824\times\frac{3000-172.6}{3000\times 172.6} = 0.097cm = \mathbf{0.97\ mm}$

δ = l_g / 2 = 0.97 / 2 ***= 0.485 mm***

5. DCresistance of the choke:

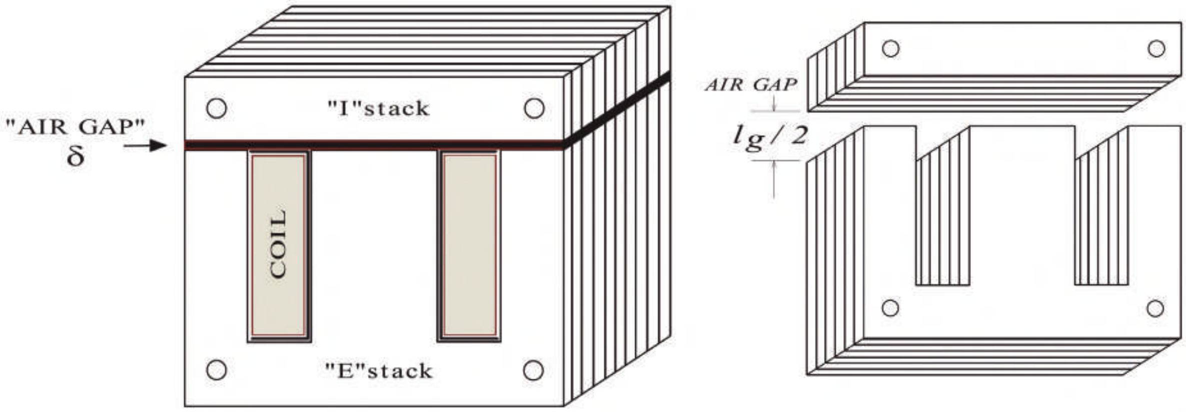
$$l_t = (2 \times b) + (2 \times h) + (\pi \times h_0) + (4 \times s) = (2 \times 3.2) + (2 \times 3.2) + (3.14 \times 1.2) + (4 \times 0.2) = 17.37\ cm$$

$$\boldsymbol{R_W} = \rho \times \frac{4 \times N \times l_t}{\pi \times d^2} = 0.0175 \times \frac{4 \times 2300 \times 0.1737}{3.14 \times 0.4^2} = \mathbf{55.6\ \Omega}$$

6. Inductance:

$$\boldsymbol{L} = \frac{1.257 \times N^2 \times S_{Fe} \times \mu_e}{l_{Fe} \times 10^8} = \frac{1.257 \times 2300^2 \times 9.535 \times 172.6}{17.824 \times 10^8} = \boldsymbol{6.17\ H}$$

Conclusion:
CLC Choke 5H / 0.25A
Manufacturing data:
Transformer laminate: EI 96, NOSS Fe 4% Si, Δ = 0.35 mm
Coil former: (3.2 cm x 3.2 cm)
Number of turns: 2250 Turns, Cul Ø 0.4 mm (insulated: Ø 0.462 *mm), 25 layer, 90 turns per layer*
Air gap: 0.35 mm (0.3 mm insulation foil + 0.05mm insulation foil)
DC Resistance: ≈ 55 Ω
Power loss: ≤ 3.5 W



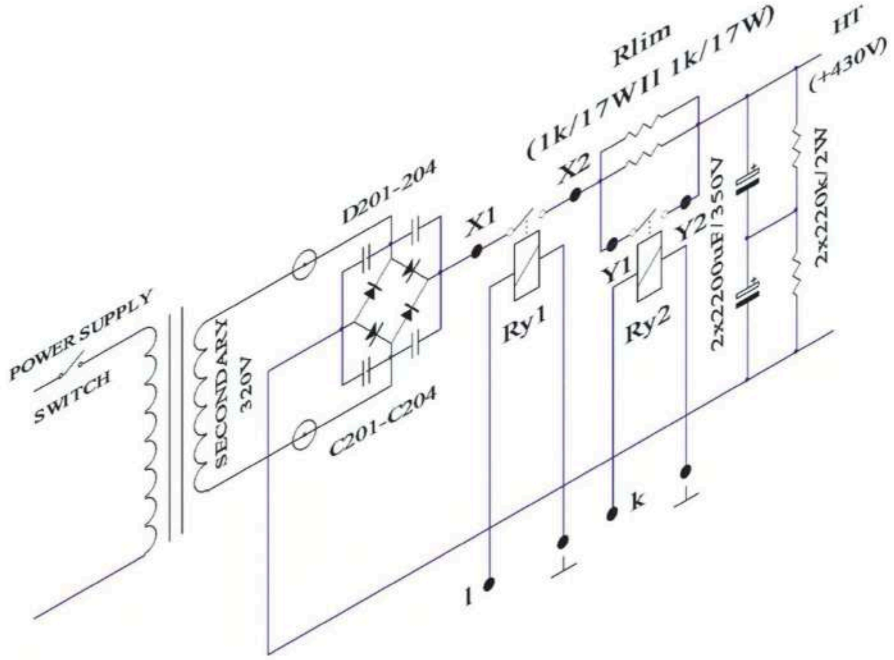
A few examples of the most commonly used Chokes

CLC Choke 10H / 0.25A
Manufacturing data:
Transformer laminate: EI 96, NOSS Fe 4% Si, Δ = 0.35 mm
Coil former: (3.2cm x 5cm)
Number of turns: 2340 Turns, Cul Ø 0.4 mm (insulated: Ø 0.462mm), 26 layer, 90 turns *per layer*
Air gap: 0.4 mm (0.4 mm insulation foil)
DC Resistance: ≈ 70 Ω
Power loss: ≤ 4.4 W

CLC Choke 10H / 0.12A
Manufacturing data:
Transformer laminate: EI 75, NOSS Fe 4% Si, Δ = 0.35mm
Coil former: (2.5cm x 2.5cm)
Number of turns: 3255 Turns, Cul Ø 0.25 mm (insulated: Ø 0.301mm), 31 layer, 105 turns per layer
Air gap: 0.25 mm (0.2 mm insulation foil + 0.05 mm insulation foil)
DC Resistance: ≈ 160 Ω
Power loss: ≤ 2.3 W

Plae Choke 100H / 36mA; f = 14 Hz
Manufacturing data:
Transformer laminate: EI 96, NOSS Fe 4% Si, Δ = 0.35mm
Coil former: (3.2cm x 4cm)
Number of turns: 7990 Turns, CuI Ø 0.2 mm (insulated: Ø 0.245 mm), 47 layer, 170 turns per layer
Air gap: 0.35 mm (0.3 mm insulation foil + 0.05 mm insulation foil)
DC Resistance: ≈ 865 Ω
Power loss: ≤ 1.12 W

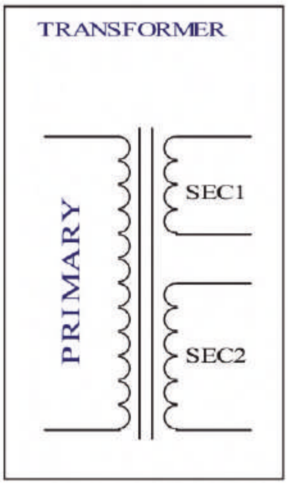
[B_{DC} = 4117 Gauss, $B_{AC}(for\ f = 14Hz) = \frac{U_{RMS}\times 10^{8}}{4.44\times N\times f\times S_{Fe}} = 16.4\times U_{RMS}$]

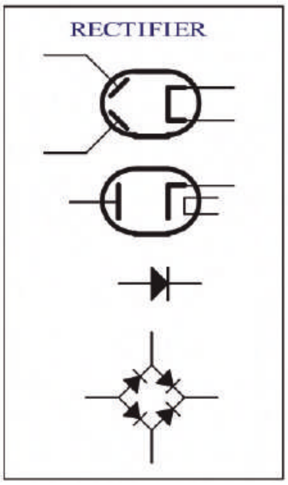

POWER SUPPLY
SWITCH
SECONDARY
320V
D201-204
C201-C204
X1
X2
Rlim
(1k/17WII 1k/17W)
HT
(+430V)
Ry1
Ry2
Y1
Y2
2x2200uF/350V
2x220k/2W
l
k

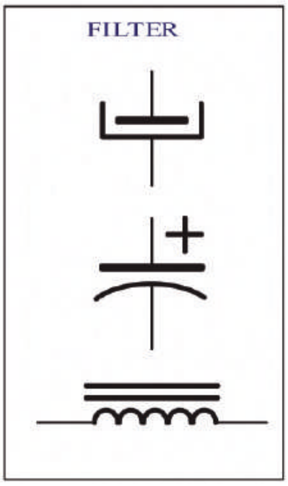
6.0 THE POWER SUPPLY

Main parts:

- **Power supply transformer** (primary — matched with standard mains voltage, secondary – one or more – according to design requirements).
- Calculation: as explained in Chapter 5.
- **Rectifier components** (rectifier vacuum tube diodes or double diodes, semiconductor diodes, rectifier diode bridges).
- **Filter components** (capacitors, inductances, resistors).
- **Auxiliary** (voltage stabilizer or voltage regulator circuit, soft start circuit, protection circuit, …)



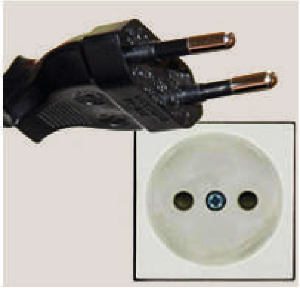




Standard mains (AC line) voltages:
Most countries use 230 V and 240 V (50 or 60 Hz) mains, and a smaller number of countries use 100 - 127 V mains.

230 V / 50 Hz (Europe, United Kingdom, India)
220 V / 50 Hz (Eastern Europe, Russian Federation, China, Far East Asia, Africa, South America)
220 V / 60 Hz (South Korea)
120 V / 60 Hz (USA, Canada, Central America, Micronesia,)
110 V / 60 Hz (Colombia, Taiwan, Virgin Islands, …)
100 V / 50 / 60 Hz (Japan)

Plugs and sockets



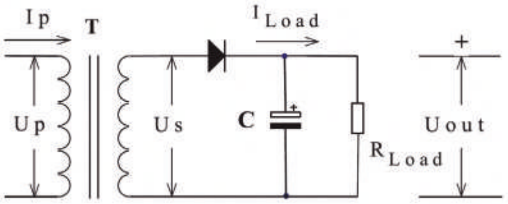
Commonly used in Europe, South America and Asia

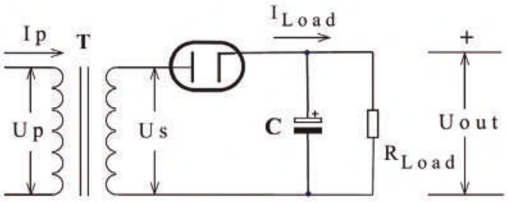
Used in Europe and Russia except UK and Ireland

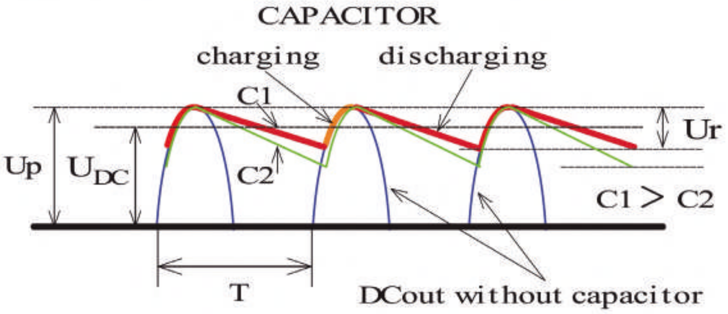
Used in UK, Ireland, Singapore, Malaysia

Used in USA, Canada, Japan

6.1 HALF - WAVE RECTIFIER (WITH CAPACITOR)







A half-wave rectifier is a very simple electronic circuit. It consists of a mains power transformer, a rectifier component (vacuum tube diode or semiconductor diode) and a filter (smoothing) capacitor.

If a filter capacitor is not used, the output current is a pulsating direct current and the rectifier circuit cannot be used to power the audio amplifier circuits.

If a filter capacitor is used, the output current of the rectifier circuit contains a DC component and an AC component (the so-called **ripple** current or voltage). Designers strive to make this unwanted DC voltage fluctuation as small as possible.

The output voltage of a half-wave rectifier circuit with a capacitor as a filter (connected in parallel to the load) is:

$$U_{DC} = \frac{2 \times f \times C \times R_L}{1 + 2 \times f \times C \times R_L} \times U_{peak}$$

$U_{peak} = 1.41 \times U_S$
U_S = *Secondary voltage RMS*
f – *mains frequency*
C – *filter capacitor*
R_L – *Load resistance*

Ripple voltage (peak to peak): $U_{r\,(p-p)} = \frac{U_{DC}}{f \times C \times R_L}$

Ripple voltage (RMS): $U_{r\,(RMS)} = \frac{U_{DC}}{2 \times \sqrt{3} \times f \times C \times R_L} = \frac{I_{Load}}{2 \times \sqrt{3} \times f \times C} = \frac{U_{peak}}{\sqrt{3}(1 + 2 \times f \times C \times R_{Load})}$

The ripple voltage is directly proportional to the load current and inversely proportional to the capacitance of the filter capacitor.

Ripple factor: $RF = \frac{U_{r\,(RMS)}}{U_{DC}} = \frac{1}{2 \times \sqrt{3} \times f \times C \times R_L}$

The mains frequency f = 50 Hz: $RF = \frac{U_{r\,(RMS)}}{U_{DC}} = \frac{1}{2 \times \sqrt{3} \times f \times C \times R_L} = \frac{5.77}{C \times R_L}$ C [μF], R [kΩ]

The mains frequency f = 60 Hz: $RF = \frac{U_{r\,(RMS)}}{U_{DC}} = \frac{1}{2 \times \sqrt{3} \times f \times C \times R_L} = \frac{4.8}{C \times R_L}$ C [μF], R [kΩ]

The ripple factor is a measure of the quality of the DC output voltage or a measure of the fluctuation of the DC voltage over time. The RF is lower if the capacitance of the filter capacitor is higher and is highly dependent on the load current. This type of power supply can be used as a power supply for circuits that do not require a high supply current for their operation (for example: fixed bias of the control grid).

Example:
Fixed bias DC voltage: – 80V, I_{Load} = 1 mA. C = 100 µF

$$RF = \frac{U_{r\,(RMS)}}{U_{DC}} = \frac{1}{2 \times \sqrt{3} \times f \times C \times R_L} = \frac{1}{2 \times \sqrt{3} \times 50 \times 100 \times 10^{-6} \times 80000} = 7.2 \times 10^{-4}$$

RF can be expressed as a percentage (%) by multiplying the RF by 100 (0.072%):

$$U_{r\,(RMS)} \text{ is: } U_{r\,(RMS)} = RF \times U_{DC} = 7.2 \times 10^{-4} \times 80 = 0.057\,V = 57\,mV$$

Ripple voltage (peak to peak):

$$U_{r\,(p-p)} = \frac{U_{DC}}{f \times C \times R_L} = \frac{80}{50 \times 100 \times 10^{-6} \times 80000} = 0.2V = 200mV$$

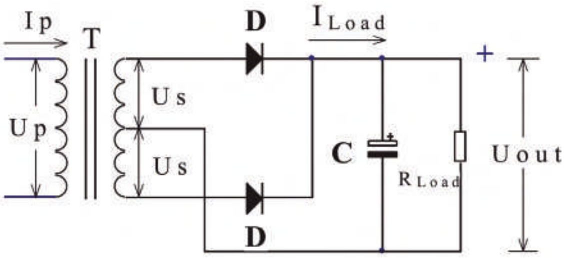
Secondary voltage is:

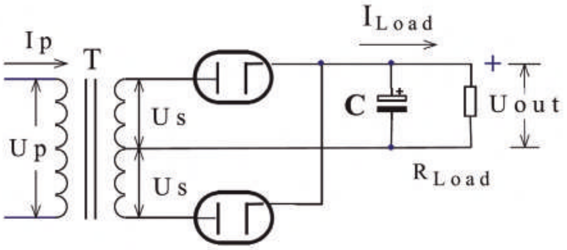
$$U_{DC} = \frac{2 \times f \times C \times R_L}{1 + 2 \times f \times C \times R_L} \times U_{peak} \rightarrow U_{peak} = \frac{1 + 2 \times f \times C \times R_L}{2 \times f \times C \times R_L} \times U_{DC}$$

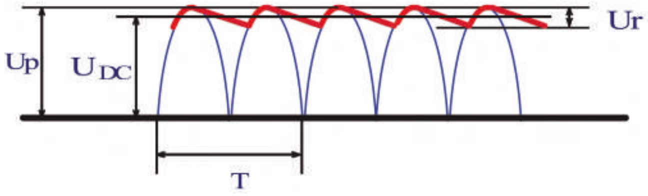
$$U_{peak} = \frac{1 + 2 \times 50 \times 100 \times 10^{-6} \times 80000}{2 \times 50 \times 100 \times 10^{-6} \times 80000} \times 80 \approx 80V \rightarrow U_{S\,(RMS)} = \frac{U_{peak}}{\sqrt{2}} = \frac{80}{1.41} = 56.7\,V$$

\# The calculation in the above example is valid for the case of an ideal rectifier. In the actual application, the voltage drop across the rectifier components and the voltage drop across the secondary winding caused by its DC resistance must be taken into account.

6.2 FULL - WAVE RECTIFIER (WITH CAPACITOR)







The output voltage of a full-wave rectifier circuit with capacitor as a filter (connected in parallel to the load) is:

$$U_{DC} = \frac{4 \times f \times C \times R_L}{1 + 4 \times f \times C \times R_L} \times U_{peak}$$

Ripple voltage (RMS): $U_{r\,(RMS)} = \frac{U_{DC}}{4 \times \sqrt{3} \times f \times C \times R_L} = \frac{I_{Load}}{4 \times \sqrt{3} \times f \times C}$

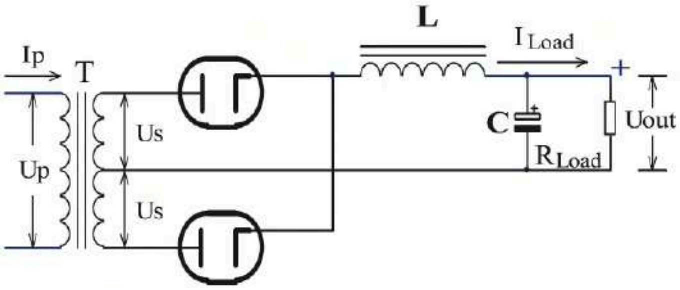
The ripple voltage is directly proportional to the load current and inversely proportional to the capacitance of the filter capacitor.

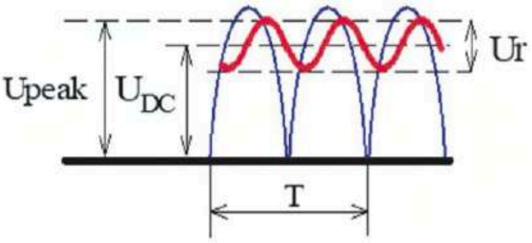
Ripple factor: $RF = \frac{U_{r\,(RMS)}}{U_{DC}} = \frac{1}{4 \times \sqrt{3} \times f \times C \times R_L}$

The mains frequency f = 50 Hz: $RF = \frac{U_{r\,(RMS)}}{U_{DC}} = \frac{1}{4\times\sqrt{3}\times f\times C\times R_L} = \frac{\mathbf{2.88}}{\boldsymbol{C\times R_L}}$ C [μF], R [kΩ]

The mains frequency f = 60 Hz: $RF = \frac{U_{r\,(RMS)}}{U_{DC}} = \frac{1}{4\times\sqrt{3}\times f\times C\times R_L} = \frac{\mathbf{2.4}}{\boldsymbol{C\times R_L}}$ C [μF], R [kΩ]

6.3 FULL-WAVE RECTIFIER WITH INPUT INDUCTANCE AND CAPACITOR AS FILTER - L FILTER





The output voltage of a full-wave rectifier circuit with inductance (connected in series) and capacitor (connected in parallel to the load), as a filter is:

$$\boldsymbol{U_{DC}} = \frac{2}{\pi}\times U_{S\,peak} = 0.637\times U_{S\,peak} = \frac{2\times\sqrt{2}}{\pi}\times U_{S\,(RMS)} = \mathbf{0.9}\times \boldsymbol{U_{S\,(RMS)}}$$

Ripple voltage (RMS):

$$U_{r\,(RMS)} = \frac{\sqrt{2}}{3}\times U_{DC}\times\frac{X_C}{X_L} = \frac{\sqrt{2}}{3}\times U_{DC}\times\frac{\frac{1}{4\times\pi\times f\times C}}{4\times\pi\times f\times L} = \frac{\sqrt{2}}{3}\times U_{DC}\times\frac{1}{16\times\pi^2\times f^2\times L\times C}$$

$$U_{r\,(RMS)} = \frac{\sqrt{2}}{48\times\pi^2\times f^2\times L\times C}\times U_{DC} = \frac{0.00298}{f^2\times L\times C}\times U_{DC}$$

The mains frequency f = 50Hz and C[μF], L[H]:

$$U_{r\,(RMS)} = \frac{1.19}{L\times C}\times U_{DC}$$

The mains frequency f = 60Hz and C[μF], L[H]:

$$U_{r\,(RMS)} = \frac{0.829}{L\times C}\times U_{DC}$$

Ripple factor:

$$RF = \frac{U_{r\,(RMS)}}{U_{DC}} = \frac{\sqrt{2}}{3}\times\frac{X_C}{X_L} = \frac{\sqrt{2}}{48\times\pi^2\times f^2}\times\frac{1}{L\times C} = \frac{0.00298}{f^2}\times\frac{1}{L\times C}$$

The mains frequency f = 50Hz and C[μF], L[H]:

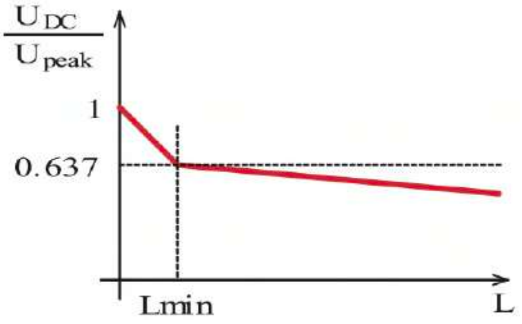
$$\boldsymbol{RF} = \frac{1.19}{L\times C}$$

The mains frequency f = 60Hz and C[μF], L[H]:

$$\boldsymbol{U_{r\,(RMS)}} = \frac{0.829}{L\times C}$$

The advantage of this type of rectifier (filter) is: **The ripple factor does not depend on the load current** (as can be seen from the equation for calculating the ripple factor). [*But this is valid only in the case when the load current is higher than some minimum value. To ensure a minimum current flow through the inductance L, some resistance (bleeder resistor) can be connected in parallel to the output of the rectifier circuit. It is common practice to set the resistance of the bleeder resistor to provide a minimum current flow equal to 1 / 10 of the load current*]. As the ripple factor does not depend on the load current, the rectifier with LC filter can be used as a power supply for audio equipment where the load current is not constant, such as class AB or class B power amplifiers.

Correct operation of this type of rectifier is ensured by using an inductance not less than the so-called **critical (minimum) inductance:**

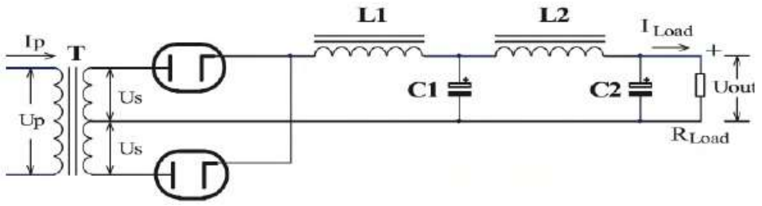


$$L_{min} = \frac{R_{Load}}{6 \times \pi \times f}$$

Therefore, for this type of rectifier, it is necessary to use a choke whose inductance is higher than the critical (minimum) inductance:

$$L \geq L_{min}.$$

Several filter cells (LC) can be connected in series:

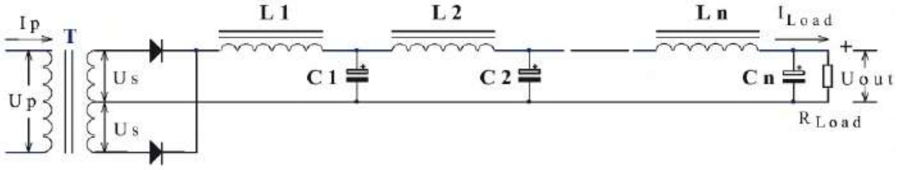


Ripple factor: $RF = \frac{\sqrt{2}}{3} \times \frac{X_{C1}}{X_{L1}} \times \frac{X_{C2}}{X_{L2}}$

For n – cells:

$$RF = \frac{\sqrt{2}}{3} \times \frac{X_{C1}}{X_{L1}} \times \frac{X_{C2}}{X_{L2}} \times \ \dots \ \times \frac{X_{Cn}}{X_{Ln}}$$

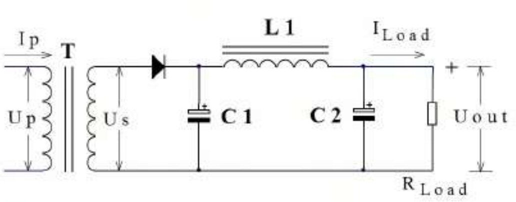
If the cells are identical:

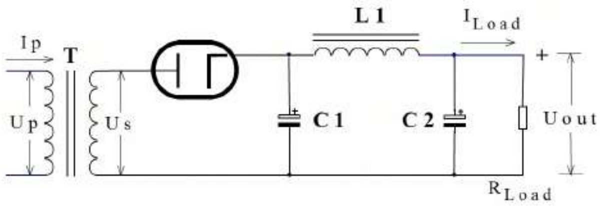


$$RF = \frac{\sqrt{2}}{3} \times \frac{1}{(8 \times \pi \times f \times L \times C)^n}$$

$C_1 = C_2 = \dots = C_n = C$;
$L_1 = L_2 = \dots = L_n = L$

6.4 HALF-WAVE RECTIFIER WITH C-L-C OR π FILTER





The output voltage of the half-wave rectifier circuit with π filter is:

$$U_{DC} = \frac{U_{S\,peak}}{1+\frac{1}{2\times f\times C_1\times R_{Load}}} \times \frac{R_{Load}}{R_{Load}+R_{choke-DC}} = \frac{\sqrt{2}\times U_{S\,(RMS)}}{1+\frac{1}{2\times f\times C_1\times R_{Load}}} \times \frac{R_{Load}}{R_{Load}+R_{choke-DC}}$$

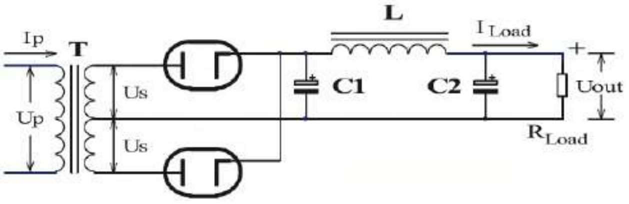
$R_{choke-DC}$ – DC resistance of the inductance L (choke)

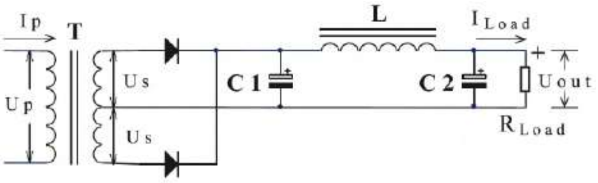
Ripple voltage (RMS):

$$U_{r\,(RMS)} = \frac{\sqrt{2}\times I_{Load}}{(2\times\pi\times f)^3\times L\times C_1\times C_2} = \frac{\sqrt{2}}{(2\times\pi\times f)^3\times L\times C_1\times C_2}\times\frac{U_{S\,peak}}{1+\frac{1}{2\times f\times C_1\times R_{Load}}}\times\frac{R_{Load}}{R_{Load}+R_{choke-DC}}$$

Ripple factor $RF = \frac{\sqrt{2}}{(2\times\pi\times f)^3\times L\times C_1\times C_2\times R_{Load}}$

6.5 FULL - WAVE RECTIFIER WITH C L C OR π FILTER





The output voltage of the full-wave rectifier circuit with π filter is:

$$U_{DC} = \frac{U_{S\,peak}}{1+\frac{1}{4\times f\times C_1\times R_{Load}}}\times\frac{R_{Load}}{R_{Load}+R_{choke-DC}} = \frac{\sqrt{2}\times U_{S\,(RMS)}}{1+\frac{1}{4\times f\times C_1\times R_{Load}}}\times\frac{R_{Load}}{R_{Load}+R_{choke-DC}}$$

$R_{choke-DC}$ – DC resistance of the inductivity L (choke)

Ripple voltage (RMS):

$$U_{r\,(RMS)} = \frac{\sqrt{2}\times U_{S\,peak}}{(4\times\pi\times f)^3\times L\times C_1\times C_2\times\left[1+\frac{1}{4\times f\times C_1\times R_{Load}}\right]\times(R_{Load}+R_{choke-DC})}$$

Ripple factor $RF = \frac{\sqrt{2}}{8\times(2\times\pi\times f)^3\times L\times C_1\times C_2\times R_{Load}}$

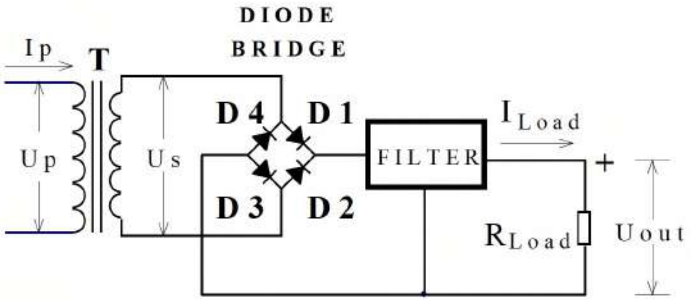
A full-wave rectifier with a π filter has **8 times less** ripple factor than a half-wave rectifier with an identical filter.

In some application where the load current is low, a resistor can be used instead of an inductance (chokes). The above equations are valid if the inductance L is substituted with R:

$$RF = \frac{\sqrt{2}}{4\times(2\times\pi\times f)^2\times R\times C_1\times C_2\times R_{Load}}$$

It is necessary to take into account the power dissipated by the filter resistor ($P_d = R \times I^2_{Load}$), because it determines the rated power of the resistor (also: the location of the resistor relative to the other electronic components of the amplifier, its cooling..). Also, the voltage drop across the filter resistor is much higher than the voltage drop across the DC resistance of the choke (it must be taken into account when calculating the voltage of the secondary winding of the power transformer).

6.6 BRIDGE RECTIFIER



The most popular full-wave rectifier circuit.
Using modern silicon diodes, this type of rectifier is widely used in almost all types of electronic equipment.

Only one secondary winding of the power transformer is required for operation of the bridge rectifier.
The voltage drop across the rectifier components (diodes) is: 2 x V_D.

All the equations used to calculate a full-wave rectifier explained above can be applied to the calculation of a bridge rectifier.

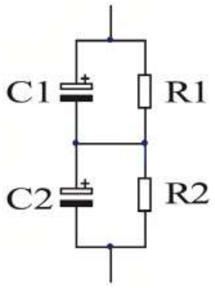
The filter (smoothing) capacitors
Aluminum electrolytic capacitors with their characteristics such as high operating voltage and high capacitance are commonly used as smoothing capacitors.

The operating voltage (not the rated voltage) of the filter capacitor is: $U_{Operating} = U_{S\,peak}$

$U_{S\ peak}$ is the peak voltage at the output of the rectifier components.

It is recommended to use a capacitors with a **rated voltage (V_R)** (Rated voltage is DC voltage value for which the capacitor is designed and specified by the manufacturer and marked on the capacitor case) 30% to 50% higher than the operating voltage of the capacitors (due to variations and peaks of the mains voltage, ambient temperature, ..).

If the rated voltage (which is standardized) is lower than the required operating voltage, two capacitors of the equal characteristics can be connected in series. Due to the tolerance of the capacitor characteristics and the possible voltage asymmetry across the capacitors, it is necessary to balance the voltages across the capacitors connected in series. A resistive voltage divider can be used to balance the voltage across the capacitors. The capacitance of an equivalent capacitor is half the capacitance of each capacitor connected in series, but the rated voltage is therefore twice the rated voltage of each capacitor connected in series.



$C_1 = C_2 = C$; $C_{eq} = C / 2$
$R_1 = R_2 = R_{Bal}$
$V_{R\ Eq} = 2 \times V_r$

The resistance of the balancing resistors is such determined to ensure that the current flowing through the resistors is approximately 20 times higher than the leakage current of the capacitors. The leakage current depends on the characteristics of the capacitor dielectric (aluminum oxide) and flows through the capacitor when a DC voltage is applied to it and also depends on the applied voltage, operating temperature …

The leakage current of a general purpose (GP) grade aluminum electrolytic capacitor can be calculated as follows:

$$I_{leak}[\mu A] = 5 \times 10^{-4} \times C[\mu F] \times U_R[V] + 3\mu A$$, U_R – rated voltage of the capacitor

Long life (LL) grade aluminum electrolytic capacitor:

$$I_{leak}[\mu A] = 2.5 \times 10^{-4} \times C[\mu F] \times U_R[V] + 1\mu A$$

Example:
C: 470μF / 350V, I_{leak} *[μA] = 5 x* 10^{-4} *x 470 [μF] x 350[V] + 3μA = 85.25μA*

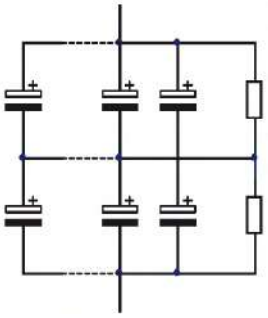
Balancing resistors for two 470μF / 350V capacitors connected in series and applied voltage of $450V_{DC}$ *:*
As the current flowing through the resistors must be approximately 20 x I_{leak} *:*
I_{Bal} *= 20 x* I_{leak} *= 20 x 85.25 =1705μA = 1.705mA.*
2 x R_{Bal} *= (applied voltage) /* I_{Bal} *= 450V / 1.7mA = 264kΩ → * R_{Bal} *= 132 kΩ*

Long life (LL) grade capacitor:
I_{leak} *[μA] = 2.5 x* 10^{-4} *x 470 [μF] x 350[V] + 1μA = 42.125μA*
I_{Bal} *= 20 x* I_{leak} *= 20 x 42.125 =842.5μA = 0.8425mA*
2 x R_{Bal} *= (applied voltage) /* I_{Bal} *= 450V / 0.8425mA = 534kΩ → * R_{Bal} *= 267 kΩ*

In practice, the resistance of a balancing resistor can be calculated using the empiric equation:

$$\boldsymbol{R_{Bal} = \frac{100}{C[\mu F]}\ [M\Omega]}$$

The above example: R_{Bal} *= 100 / 470* [μF] *= 0.212 MΩ = 212 kΩ*

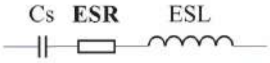


When higher capacitance is required, several pairs of capacitors connected in series can be connected in parallel.

$$R_{Bal} = \frac{100}{C[\mu F] \times \sqrt{n}} \, [M\Omega]$$

n - number of parallel-connected pairs of series-connected capacitors

ESR (Equivalent series resistance)



There is no ideal capacitor (pure capacitance). The equivalent circuit of an actual electrolytic capacitor consists of several components connected in series:

- C_s (capacitance) – reactance of capacitance: 1 / (2 x π x f x C_s)
- ESR (equivalent series resistance) – dielectric losses and resistance of the electrolyte, foils and the terminals.
- ESL (equivalent series inductance) – reactance of inductance: 2 x π x f x ESL (electrolyte, foils winding and the terminals).

It is important that the equivalent impedance of the capacitor is as low as possible due to its role in a situation where the load resistance of the power supply changes very quickly and abruptly to a lower value (surge current – *"a current that increases or decreases from the normal rated value for a short time"*). The resistance of the secondary winding of the power transformer and rectifier components can be relatively high and abruptly surge currents can cause a voltage drop across them and a voltage drop across the load. This situation can happen in a power amplifier because the signal in a power amplifier represents music that can contain a lot of transients (sudden high-level sound produced by drums, trumpet, piano, opera vocals…). (more noticeable in class B and AB amplifiers than in class A amplifier). Variations in the output voltage of the rectifier stage can cause unwanted distortion of the power amplifier signal. In such a situation, the lack of energy is compensated by the energy stored by the filter capacitors (especially by the electrical energy stored by the last capacitor of the power supply chain).

The energy stored by the capacitor is:

$\boldsymbol{J = \frac{1}{2} \times C \times U^2}$; J – energy [Joules], C – capacitance [Farads], U – voltage [Volts]

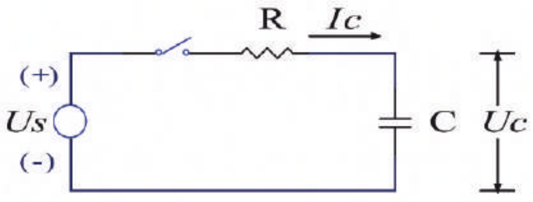
Therefore, it is useful to use capacitors with the highest possible capacitance (especially the last capacitor of the power supply chain) taking into account the rated values of the other components used in a rectifier circuit (transformer, rectifier components…).

The impedance of the filter electrolytic capacitor (C_s – ESR – ESL) affects the total output impedance of the power supply circuit and its frequency response (especially at high frequencies). To reduce this effect, non-polarized capacitors (like polypropylene, Teflon…) can be connected in parallel with filter electrolytic capacitors.

The use of high capacitance capacitors (especially the first filter capacitor on the back side of the rectifier components) can cause problems and damage other components of the power supply (rectifier components and transformer) especially at the moment of turning on the power supply. The capacitor charging current can be very high and the power and current limits of the rectifier components can be exceeded.

The charging current can be limited by using a resistor in the capacitor charging circuit. The resistance of the capacitor charging current limiting resistor must be chosen so that the charging current is lower than the rated current of the rectifier diodes (it is not necessary to oversize the rated power of the resistor because the charging time of the capacitor is not so long).

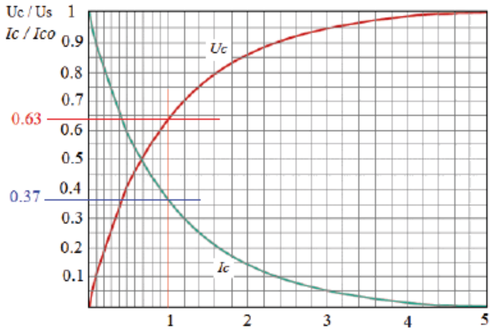
Capacitor charging:



Uc – instantaneous voltage across the capacitor:

- $U_c = U_s(1 - e^{-\frac{t}{\tau}})$

Us – source voltage
τ – Time constant



τ = R[Ω] x C[F]

I_c – instantaneous capacitor charging current:

$I_c = I_{C0} \times e^{-\frac{t}{\tau}}$

I_{C0} = *Us / R*

t – elapsed time from the moment of applying voltage to the capacitor

(t)

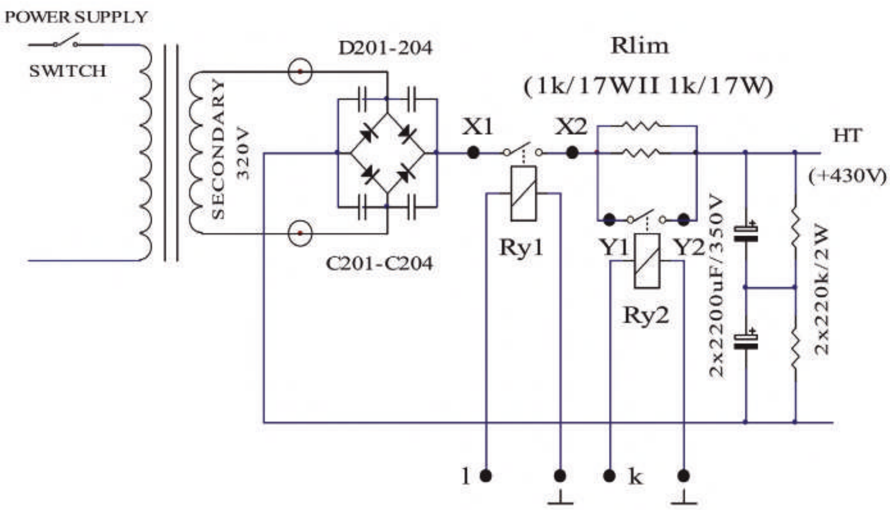
The above equations can be applied and used in practice. For example, the capacitor C can be charged by a DC voltage source Us via the series resistor R to a voltage V_c in the time period t = τ = R × C:

$$U_c = U_s\left(1 - e^{-\frac{t}{\tau}}\right) = U_s(1 - 2.718^{-1}) = 0.629 \times U_s$$

The voltage across the capacitor after the time t = R × C elapsed from the moment of applying the voltage to the capacitor is 0.629 × Us.
After the time period t = 5 x τ the capacitor will be almost fully charged – the voltage across capacitor will be almost equal to the source voltage ($U_C = 0.98 \times U_S$).
The maximum charging current is at the moment when the source voltage is applied to the capacitor: $I_{c0} = U_S / R$.
After time t = τ the charging current is:

$$I_c = I_{C0} \times e^{-\frac{t}{\tau}} = \frac{U_S}{R} \times \frac{1}{2.718} = 0.368 \times \frac{U_S}{R}$$

Example:
High voltage power supply:



Here, A bridge rectifier is used.
Rated diode current: 3 A.
Secondary voltage: 320 V_{RMS}
Filter capacitor: two 2200 μF capacitors connected in series.

Project requirement:
The capacitor charging current must be lower than 1 A.
Calculation of R and the time required for the voltage across the capacitor to reach 63% of the voltage at the output of the rectifier bridge.

Calculation:
The peak voltage of the bridge rectifier is: $U_S = \sqrt{2} \times U_{S\,[RMS]} = \sqrt{2} \times 320 = 451.2V$

Maximum charging current (power - on moment): $I_{c0} = \frac{U_S}{R} \rightarrow R = \frac{U_S}{I_{c0}} = \frac{451.2}{1} = 451.2\Omega$

To make the calculation easier, let's take (for example) R = 500Ω → $I_{c0} = 451.2/500 = 0.9A < 1A$

The time constant is: $\tau = R \times C = 500 \times 1100 \times 10^{-6} = 0.55$ *s.*
(C = two 2200μF capacitors connected in series = 1100μF).
Using the diagram above (Uc/Us = f (t)):
After the time period t = 1 x τ = 0.55s, the voltage across the capacitor is 0.63 x $U_S = 0.63 \times 451.2 = 285V$.

The charging current at that time is: $I_c = 0.368 \times \frac{U_S}{R} = 0.368 \times \frac{451.2}{500} = 0.332A$

0.55s after the power supply is turn on, the voltage across the capacitor is 285V, the resistor R can be short-circuited, and the process of charging the capacitor can continue up to voltage U_S.
In actual application, the voltage drop on the rectifier diodes and the voltage drop on the primary and secondary windings of the power transformer must be taken into account.
The DC voltage at the output of the power supply circuit is:

$$U_{out} = \sqrt{2} \times U_{S[RMS]} - I_{Load} \times \left(R_{S[DC]} + n^2 \times R_{P[DC]}\right) - 2 \times U_D$$

$R_{S[DC]}$ – *DC resistance of the secondary winding*
$R_{P[DC]}$ – *DC resistance of the primary winding*

n – Turns ratio, $n = \frac{N_S}{N_P}$;

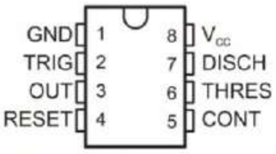
N_S – *number of turns of the secondary winding,* N_P – *number of turns of the primary winding*
$n^2 \times R_{P[DC]}$ *is the resistance of the primary winding reflected to the secondary.*

Practical delay circuit
The above list of high voltage power supply design requirements is extended by one more requirement:
The DC high voltage (HT) output of the power supply must be activated after 120 seconds (approximately) from the moment the amplifier is turned on. This time (120s) is sufficient for the cathodes of the power amplifier tubes to heat up to an acceptable level when the anode voltage can be applied.

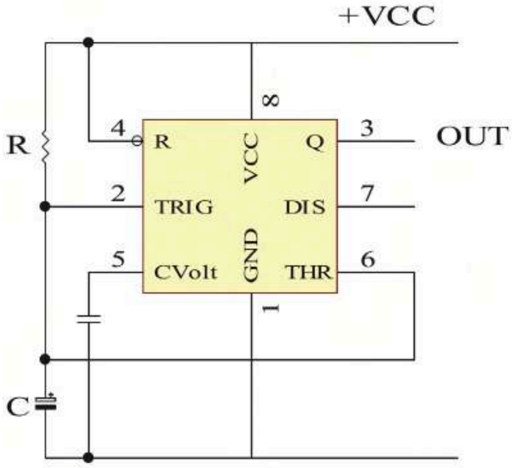
1. *The first relay Ry1 must be activated 120 seconds after the power supply is turned on and remains ON as long as the amplifier is running (until the power amplifier is turned off).*
   (T_1 *= 120 seconds - time delay).*

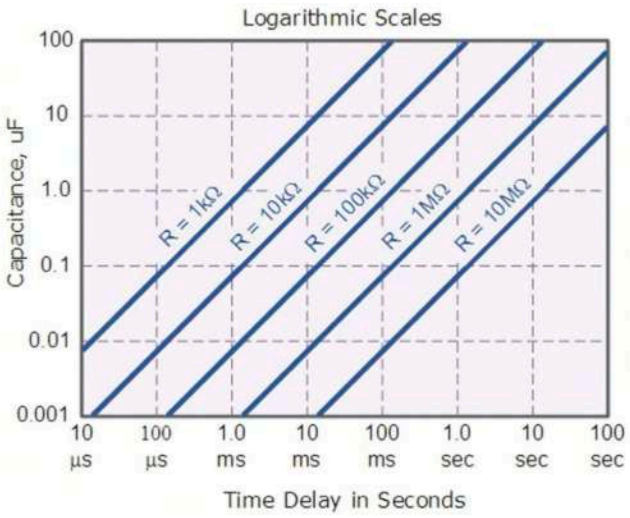
2. *The second relay Ry2 must be activated 0.55 second after the moment of activation of the first relay and remains ON as long as the amplifier is running (until the power amplifier is turned off).*
   (T_2 *= 0.55 seconds - time delay).*
   During a time period of 0.55 second the capacitors are charged via a high power current limiting resistors R_{Lim}*. After this time period, the second relay is activated and its contacts short-circuit the limiting resistor and the process of charging the capacitors continues via the closed relay contacts until the capacitors are fully charged.*

The well-known IC NE555 timer can be used to make a time delay circuit.



Basic Long Time Delay Circuit Configuration:





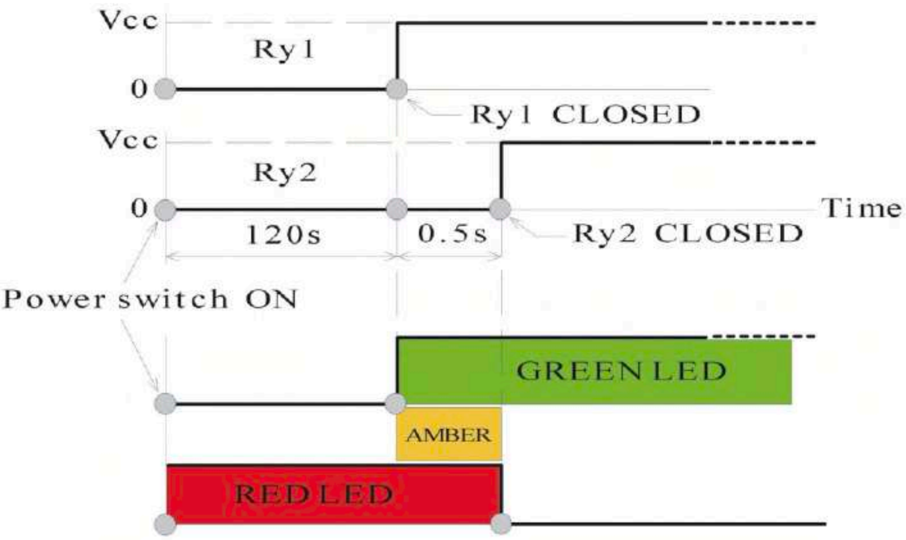
The time delay constant is:

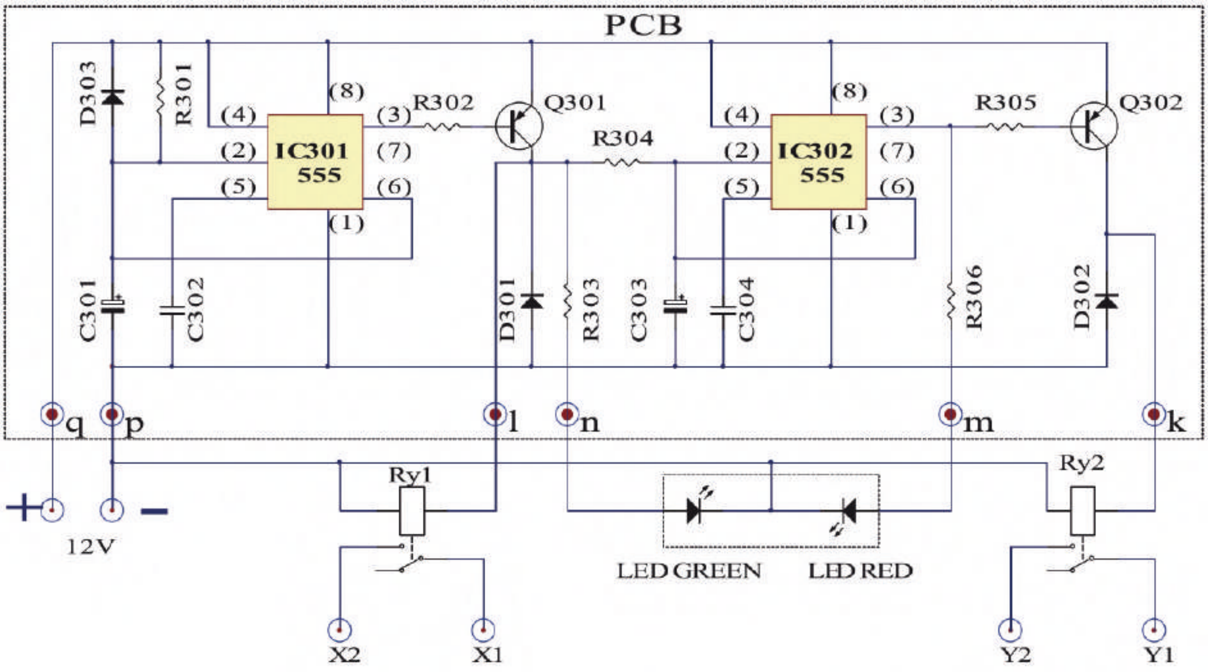
$$T = 1.1 \times R \times C$$

T_1 = 120 s
R_1 = 1MΩ
C_1 = T / (1.1 x R_1) = 120 / (1.1 x 1MΩ) = 109 μF
C_1 = 100 μF

T_2 = 0.55 s
R_1 = 1MΩ
C_2 = T / (1.1 x R_2) = 0.55 / (1.1 x 1MΩ) = 0.5 μF
C_2 = 0.47 μF

Time diagram





PCB – 3xx

R301 – 1M
R302 – 10k
R303 – 1k
R304 – 1M
R305 – 10k
R306 – 1k

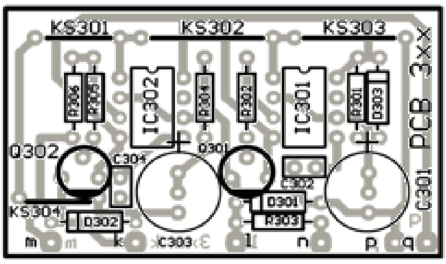
C301 – 100 μF/16V
C302 – 100 nF/63V
C303 – 0.47μF/16V
C304 – 100 nF/63V

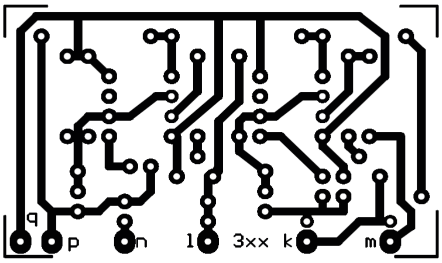
D301 – 1N4148
D302 – 1N4148
D303 – 1N4148

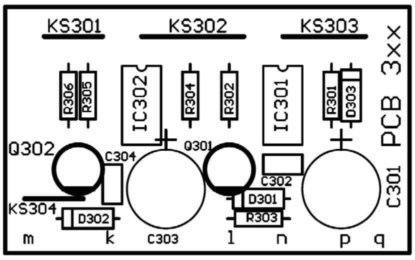
Q301 – BC 557B
Q302 – BC 557B

IC301 – NE555
IC302 – NE555

Ry1 – 12V (250V/15A)
Ry2 – 12V (250V/15A)
LED – RED - GREEN







Notes on the electronic components of the high-voltage power supply

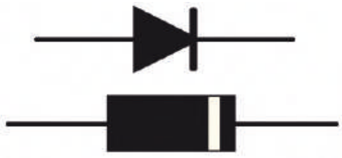
Filter capacitors

Maximum permissible operating temperature — the maximum permissible ambient temperature at which the capacitor can operate continuously (usually 85 ^{0}C and 105 ^{0}C).

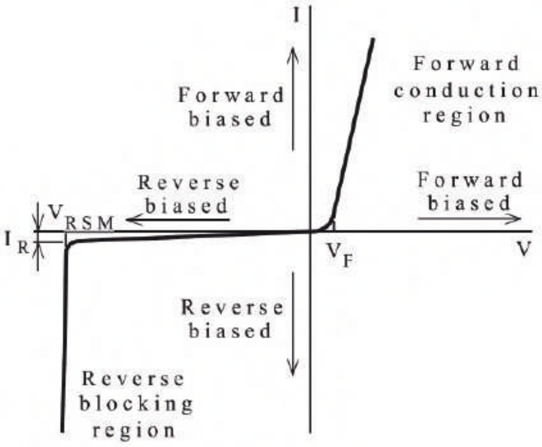
Therefore, it is necessary to place the capacitors in places inside the device where the capacitors are not exposed to the high temperature of components that emit heat such as tubes, power resistors, heat sinks.

The next note refers to the **age** of the capacitor. Branded capacitor manufacturers print the production date on the capacitors (four-digit date code: **MM.YY**, month and year of manufacture). If the capacitor is stored for a long period of time (several years), it needs to be reformed before it is installed in the device. A simple procedure for reforming the capacitor is: connect the capacitor to a DC power supply of the rated voltage of the capacitor via a series resistor of 100 Ω for $V_R \leq 100\ V_{DC}$ and 1000 Ω for $V_R \geq 100\ V_{DC}$ for a period of one hour. After that, the capacitor is ready for use.

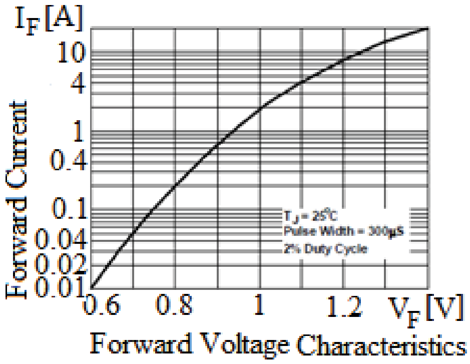
About **Rectifier Diodes (Semiconductor)**



A typical U – I characteristic of a silicon rectifier diode:



V_F (Forward voltage)*:*



This is the voltage dropped across a conducting, forward-biased diode.
Silicon diode (typical): V_F = (0.6V ... 1.4V)
Silicon Schottky diode : V_F = (0.2V ... 0.3V).

Illustration: Forward voltage characteristics ($I_F = f(V_F)$) of a silicon rectifier diode (1N4007).

$I_{F(AV)}$ *(***Average rectified forward current):**
The highest current in continuous operation, averaged over one complete cycle of the operating frequency.

Never use diode operation at $I_{F(AV)}$ in practice, especially if the diodes are mounted on a PCB – the safe operating current of the diode is about half $I_{F(AV)}$.

V_R (Maximum DC Reverse Voltage) and V_{RRM} (Maximum Repetitive Reverse Voltage):
V_R - Maximum continuous voltage the diode can withstand in reverse biased mode on a continual basis..
V_{RRM} - Maximum voltage the diode can withstand in reverse biased mode, in repeated puses.
Half-wave and full-wave (two diodes and a center-tapped transformer) rectifier configuration:

$$V_{RRM} > \sqrt{2} \times 2 \times U_S = 2.83 \times U_S;$$

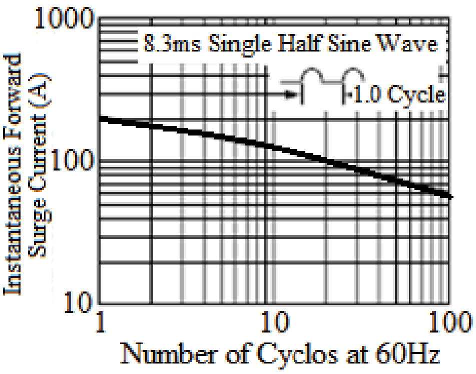
U_S = Secondary voltage (RMS).

Bridge rectifier configuration:

$$V_{RRM} > \sqrt{2} \times U_S = 1.41 \times U_S$$

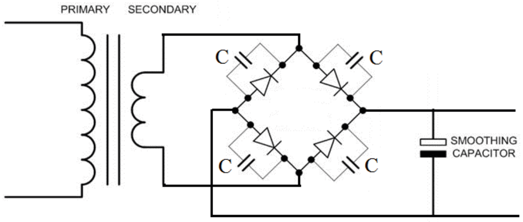
It is essential to fulfill the above condition.

I_{FSM} (Non – repetitive Peak Forward Current:
The maximum non-repetitive on-state surge current is generally quoted for one 10 (or 8.6) milliseconds sinusoidal period (Single Half – Sine – Wave). Such an overload situation can be tolerated only a limited number of times before failure results.
10 ms corresponds to one half cycle of (50Hz) mains
8.6 ms corresponds to one half cycle of (60Hz) mains
I_{FSM} characteristics of a typical silicon rectifier diode (1N5408):



t_{rr} *(Reverse Recovery Time):*
A measure of the time taken for a rectifier diode to revert to a blocking state after the voltage across it has been reversed.

One more note:
To protect the power supply diodes from externally generated spikes reflected from the mains to the transformer secondary is to connect a capacitance (capacitor) in parallel to each of the four diodes in the bridge rectifier.
(C = [0.01 ... 0.1] µF)

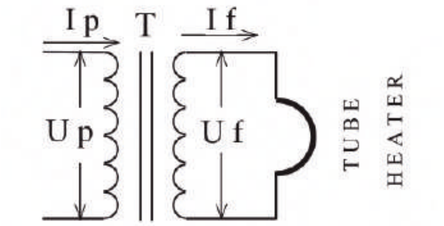


Summary:
The characteristics of a quality power supply are:

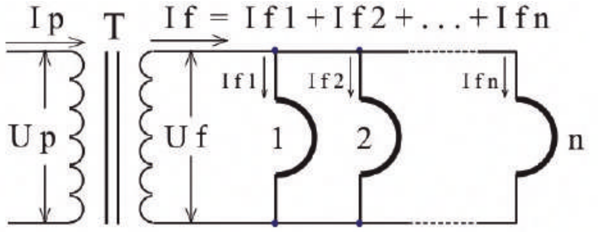
- Low output impedance — frequency independent (good linearity)
- Large capacity of stored electrical energy
- Minimum variation of the output voltage
- Minimum ripple factor and minimum hum and noise

Tube heater power supply
AC power supply

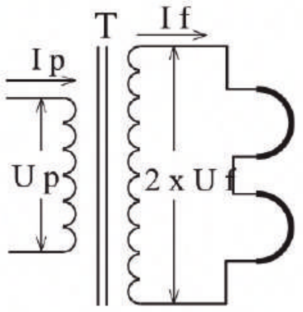
In practice, a power transformer with a secondary voltage equal to the rated AC voltage U_f of the tube heater specified by the manufacturer (mostly standardized, 6.3V, 4V, 5V, 12.6V…) is most often used. Two or more tubes (tube heaters or heaters of each section of multi-section tubes) can be connected in parallel or in series and can be powered from the common secondary winding of the power transformer. The power of the transformer secondary must be calculated according to the power of the heaters it supplies.



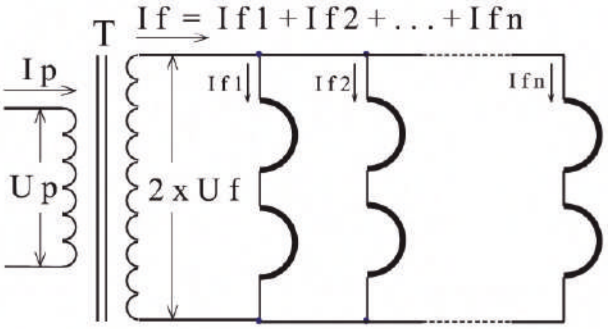
Power of the Secondary $P_{S(f)} = I_f \times U_f$



$P_{S(f)} = I_{f1} \times U_f + I_{f2} \times U_f + \cdots + I_{fn} \times U_f$



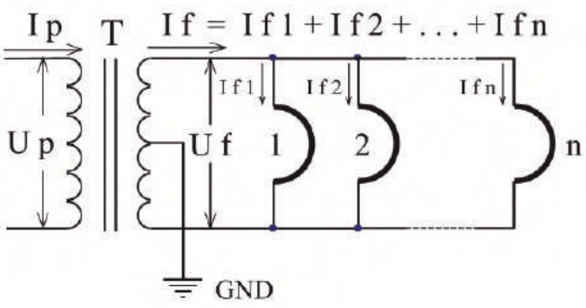
Power of the Secondary $P_{S(f)} = I_f \times 2 \times U_f$

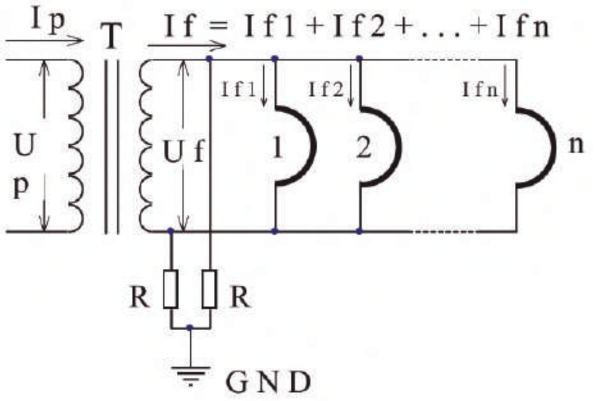


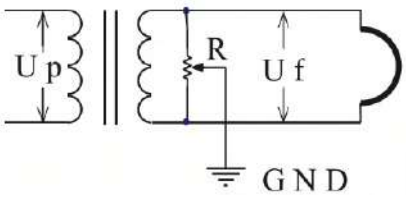
$P_{S(f)} = 2 \times U_f \times (I_{f1} + I_{f2} \times + \cdots + I_{fn})$

KT88: 6.3V, 1.6A (power of the secondary: $P_{S(f)}$ = 6.3V x 1.6A = 10.08W)
Two KT88 (heaters connected in parallel): 6.3V, 3.2A ($P_{S(f)}$ = 20.16W)
Two KT88 (heaters connected in series): 12.6V, 1.6A ($P_{S(f)}$ = 20.16W)

The secondary winding of the power transformer for supplying the tube heater can be made with a central tap (CT). The central tap can be earthed (grounded). If the secondary does not have a central tap, it can be simulated and made using two resistors (R = 100Ω ÷220 Ω) in a series connection connected in parallel with the secondary winding. The common junction of two resistors simulates the central tap of the secondary winding which can be grounded.

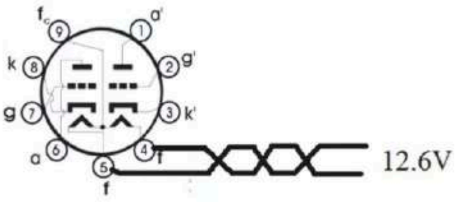


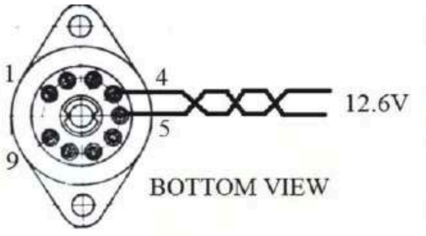


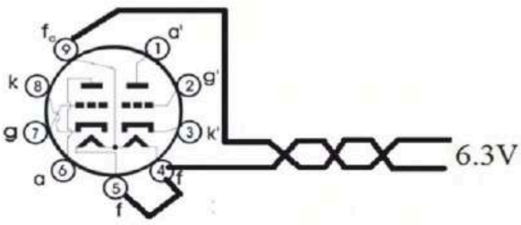


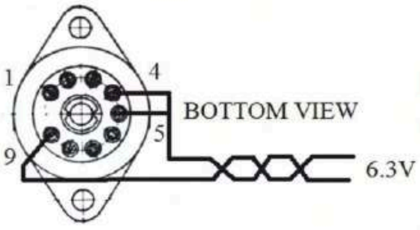
Instead of two fixed resistor R, a variable wire wound resistor can be used to balance the heater circuit in order to minimize hum and noise (this method is very efficient for powering directly heated tubes).

Double low-power ECC type triodes (ECC81, ECC82, …). Each triode section has its own heater (pin 4 – pin 9 and pin 5 – pin 9). These two heaters share one common connection point (pin 9), so, they can be supplied by one 6.3V, 0.3A secondary winding (two heaters in parallel connection) or by 12.6V, 0.15A secondary winding (two heaters in series connection).





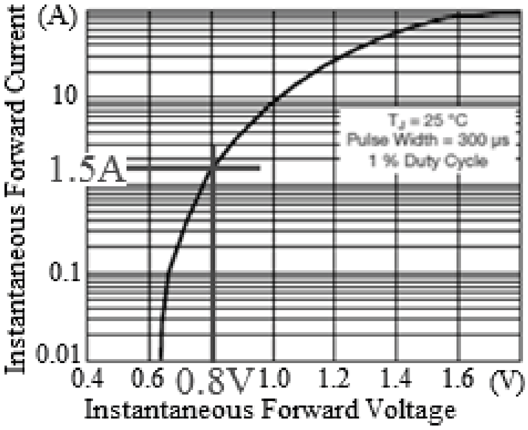




The heater power supply wiring must be made using wires of the appropriate diameter.
(Wire cross-section [mm^2] = I_f [A] / (3 … 4)). (Wire diameter [mm] = (0.56 … 0.65) x square root (I_f [A])).

Due to the relatively high AC current (mains frequency) flowing through the wires connecting the secondary of the power transformer to the tube heaters (heater socket pins), the generated electromagnetic field can have a harmful effect on the amplifier signal (hum and noise). In order to minimize this problem, it is necessary to use a pair of twisted wires.

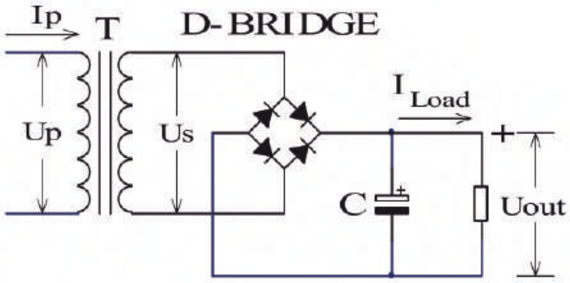
DC heater power supply



Bridge rectifier with filter capacitor
Secondary winding of the transformer: 6.3V AC, 1.5A;
Mains frequency: f = 50 Hz

Project requirements:
$U_f = 6.3V$ *DC /* $I_f = 1.5A$ *DC,*
$R_{Load} = U_f / I_f = 6.3 / 1.5 = 4.2\ \Omega$
Ripple factor: RF ≤ 10%

Design:
Diode: $I_{F(AV)} > 1.5A$, $V_{RM} > 6.3V$
The 1N5402 Si diode fulfills the above requirements
1N5402: $I_{F(AV)} = 3A$, $V_{RRM} = 200V$
Forward voltage: $V_F = 0.8V$ *at* $I_{F(AV)} = 1.5A$

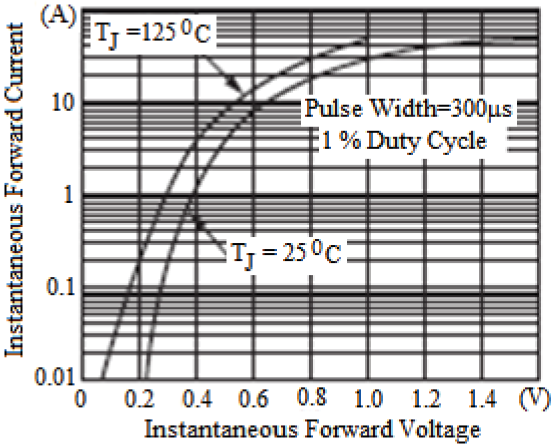


$$\boldsymbol{RF} \le \frac{U_{r\,(RMS)}}{U_{DC}} = \frac{2.88}{C[\mu F] \times R_L[k\Omega]} = \frac{2880}{C[\mu F] \times R_L[\Omega]} \le 0.1 \rightarrow C[\mu F] \ge \frac{2880}{0.1 \times R_L[\Omega]} = \frac{2880}{0.1 \times 4.2]} = \mathbf{6857\mu F}$$

Standard, C = 10 000μF / 16V

$$RF = \frac{U_{r\,(RMS)}}{U_{DC}} = \frac{2880}{C[\mu F] \times R_L[\Omega]} = \frac{2880}{10000 \times 4.2} = \mathbf{0.068}\ \ \boldsymbol{or}\ \ \mathbf{6.8\%}$$ *or, ripple voltage: 0.068 x 6.3V = 0.428V = 428mV*

$$U_{DC} = \frac{4 \times f \times C \times R_L}{1 + 4 \times f \times C \times R_L} \times U_{peak} - 2 \times V_F = \frac{4 \times 50 \times 10^{-2} \times 4.2}{1 + 4 \times 50 \times 10^{-2} \times 4.2} \times \sqrt{2} \times 6.3 - 2 \times 0.8 = 6.3V$$

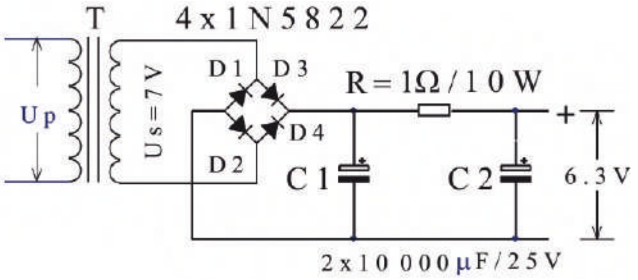


Notes:
The power dissipation of the 1N5402 diode is $P_d = I_{F(AV)} \times V_{RM} = 1.5 \times 0.8V = 1.2W$, and therefore they can get very hot during operation, especially if mounted on a PCB.

Schottky diodes such as the 1N582x type diodes are a better choice than the universal Si rectifier diodes.

1N582x *: $I_{F(AV)}$ = 3A, V_F: 0.475V ÷ 0.525V*
V_{RRM} = 20V – 1N5820,
V_{RRM} = 30V – 1N5821,
V_{RRM} = 40V – 1N5822

Improved version: bridge rectifier with CRC filter: 6.3V, 1.5A



Ripple factor

$$RF = \frac{\sqrt{2}}{4 \times (2 \times \pi \times f)^2 \times R \times C_1 \times C_2 \times R_{Load}}$$

$\boldsymbol{RF} = 0.85\%$ *or,*

Ripple voltage: $U_r = 0.0085 \times 6.3\ U_{DC} = 53.55$ mV

$$U_{DC} = \frac{4 \times f \times C \times R_L}{1 + 4 \times f \times C \times R_L} \times U_{peak} - 2 \times V_F - R \times I_{Load}$$

$$\boldsymbol{U_{DC}} = \frac{4 \times 50 \times 10^{-2} \times 4.2}{1 + 4 \times 50 \times 10^{-2} \times 4.2} \times \sqrt{2} \times 7 - 2 \times 0.5 - 1 \times 1.5 = \mathbf{6.32V}$$

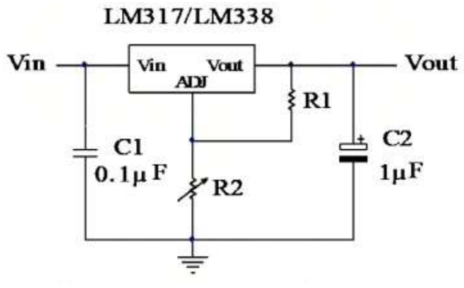
Secondary winding of the transformer: $U_S = 7V_{RMS}$
D1, D2, D3, D4 – 1N5822 (3A, 40V, Schottky)
C1, C2 – 10 000µF / 16V
R - 1Ω / 10W

U_{DC} = 12.6V *heater power supply. The secondary voltage of the power transformer is:*

$$\boldsymbol{U_S} = (\boldsymbol{U_{DC}} + 2 \times V_F + R \times I_{Load}) \times \frac{1 + 4 \times 50 \times 10^{-2} \times 8.4}{4 \times 50 \times 10^{-2} \times 8.4} \times \frac{1}{\sqrt{2}} = \mathbf{11.35V} \approx \mathbf{12.6V}$$

** To adjust the output voltage of 12.6V, use the power resistor R (1.5Ω ... 1.8Ω).*

Regulated DC heater power supply
To keep the heater supply voltage stable and to minimize ripple voltage (hum and noise in the amplifier caused by insufficiently filtering of the power supply), a regulated DC heater power supply circuit can be used.
Two types of adjustable linear IC regulators are commonly used to make a heater power supply circuit:
LM317 adjustable 3 - terminal positive voltage regulator capable of supplying 1.5A in the range of 1.2V to 37V output voltage and LM338 capable of supplying 5A in the range of 1.2V to 32V output voltage.



Basic configuration of the regulated power supply circuit:

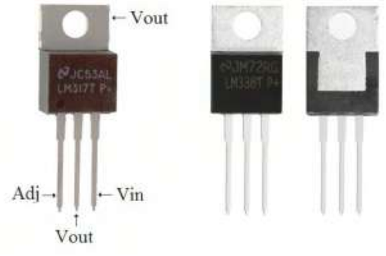
A few notes about the LM 317 and LM 338 IC regulators:

- *Minimum difference between input and output voltage of the IC:* $U_{in\ (min)} - U_{out} > 3V$.

- *Power dissipation of the IC:* $(U_{in} - U_{out}) \times I_{Load} < 15W$.

Output voltage calculation:

$$U_{out} = 1.25 \times \left(1 + \frac{R_2}{R_1}\right) + I_{Adj} \times R_2$$



$I_{Adj} = (40 \div 100)\ \mu A$
R_1 – 240 Ω (LM317)
R_1 – 120 Ω (LM338)

$(U_{in} - U_{out}) > 3V \rightarrow U_{in\ (min)} > U_{out} + 3V$
$(U_{in} - U_{out}) \times I_{Load} < 15W$

Example
Regulated DC heater power supply

1. $U_{out} =$ 6.3V DC, ($I_{Load} <$ 1.5A, LM 317), ($I_{Load} <$ 2.5A, LM 338)
2. $U_{out} =$ 12.6V DC, ($I_{Load} <$ 1.5A, LM 317), ($I_{Load} <$ 2.5A, LM 338)

Calculation of ***R_2*** *resistor:*

LM 317 (R_1 = 220 Ω) and U_{out} = 6.3V:

$$U_{out} = 1.25 \times \left(1 + \frac{R_2}{R_1}\right) + I_{Adj} \times R_2 \rightarrow R_2 \approx \left(\frac{U_{out}}{1.25} - 1\right) \times R_1 = \left(\frac{6.3}{1.25} - 1\right) \times 220 \approx 889\Omega$$

LM 338 (R_1 = 120 Ω) and U_{out} = 6.3V:

$$U_{out} = 1.25 \times \left(1 + \frac{R_2}{R_1}\right) + I_{Adj} \times R_2 \rightarrow R_2 \approx \left(\frac{U_{out}}{1.25} - 1\right) \times R_1 = \left(\frac{6.3}{1.25} - 1\right) \times 120 \approx 485\Omega$$

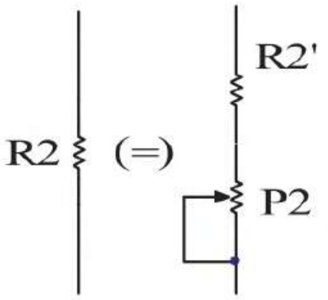
LM 317 (R_1 = 220 Ω) and U_{out} = 12.6V:

$$U_{out} = 1.25 \times \left(1 + \frac{R_2}{R_1}\right) + I_{Adj} \times R_2 \rightarrow R_2 \approx \left(\frac{U_{out}}{1.25} - 1\right) \times R_1 = \left(\frac{12.6}{1.25} - 1\right) \times 220 \approx 1997\Omega$$

LM 338 (R_1 = 120 Ω) and U_{out} = 12.6V:

$$U_{out} = 1.25 \times \left(1 + \frac{R_2}{R_1}\right) + I_{Adj} \times R_2 \rightarrow R_2 \approx \left(\frac{U_{out}}{1.25} - 1\right) \times R_1 = \left(\frac{12.6}{1.25} - 1\right) \times 120 \approx 1090\Omega$$

To allow fine adjusting of the output voltage, R_2 *can be replaced by a series connection of a fixed resistor* (R_2') *and a variable trimmer potentiometer* (P_2)*:*



	LM317 (R_1 = 220 Ω)	LM338 (R_1 = 120 Ω)
U_{out} = 6.3 V	R_2' = 680 Ω	R_2' = 270 Ω
U_{out} = 12.6 V	R_2' = 1k8	R_2' = 820 Ω
P_2 = 470 Ω		

Minimum voltage $\boldsymbol{U_{S\,(min)}}$ *of the secondary winding:*

- *The minimum input voltage of the voltage regulator is:* $U_{in\,(min)} > U_{out} + 3V$
- $U_{out} = 6.3\ V$: $U_{in\,(min)} > 6.3V + 3V = 9.3\ V$
- $U_{out} = 12.6\ V$: $U_{in\,(min)} > 12.6V + 3V = 15.6\ V$

If a bridge rectifier with a CRC filter is used, as in the example above, the voltage of the secondary winding is:

$$U_S = \left(U_{in\,(\min)} + 2 \times V_F + R \times I_{Load}\right) \times \frac{1 + 4 \times f \times C \times R_L}{4 \times f \times C \times R_L} \times \frac{1}{\sqrt{2}}$$

I_{Load} *= 1.5A (LM317), C = 10 000µF, R = 1Ω ,* $R_L = U_{OUT} / I_{Load}$ *(4.2Ω for 6.3V and 8.4Ω for 12.6V) and Schottky diode 1N5822 (*$V_F \approx 0.5V$*):*

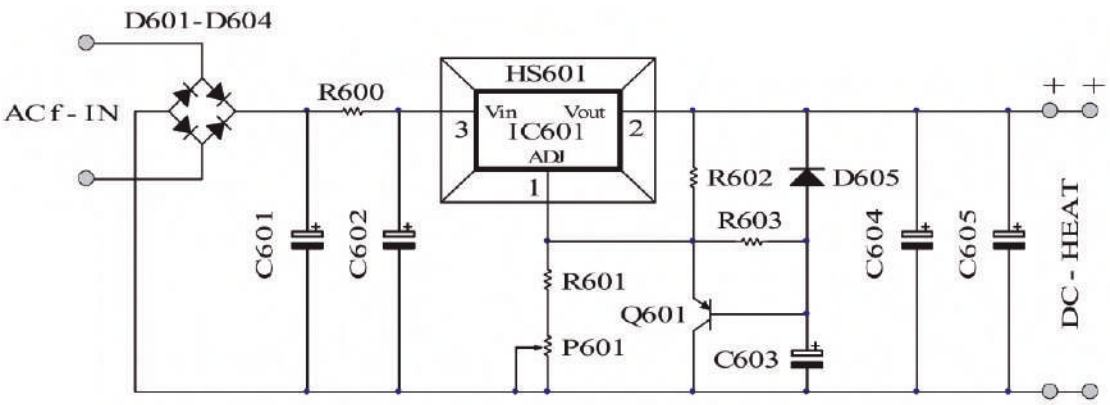
$$6.3V\ /\ 1.5A\!:\ U_S > (9.3 + 2 \times 0.5 + 1 \times 1.5) \times \frac{1{+}4{\times}50{\times}10^{-2}{\times}4.2}{4{\times}50{\times}10^{-2}{\times}4.2} \times \frac{1}{\sqrt{2}} = 9.36V \quad (\textit{The choice: } \boldsymbol{U_S = 10V_{RMS}})$$

$$12.6V\ /\ 1.5A\!:\ U_S > (15.6 + 2 \times 0.5 + 1 \times 1.5) \times \frac{1{+}4{\times}50{\times}10^{-2}{\times}8.4}{4{\times}50{\times}10^{-2}{\times}8.4} \times \frac{1}{\sqrt{2}} = 13.6V \quad (\textit{The choice: } \boldsymbol{U_S = 15V_{RMS}})$$

I_{Load} *=2.5A (LM338), C = 10 000µF, R = 0.5Ω ,* $R_L = U_{OUT} / I_{Load}$ *(2.52Ω for 6.3V and 5.04Ω for 12.6V) and Schottky diode 1N5822 (*$V_F \approx 0.5V$*):*

$$6.3V\ /\ 2.5A\!:\ U_S > (9.3 + 2 \times 0.5 + 0.5 \times 2.5) \times \frac{1{+}4{\times}50{\times}10^{-2}{\times}2.52}{4{\times}50{\times}10^{-2}{\times}2.52} \times \frac{1}{\sqrt{2}} = 9.8V \quad (\textit{The choice: } \boldsymbol{U_S = 10V_{RMS}})$$

$$12.6V\ /\ 2.5A\!:\ U_S > (15.6 + 2 \times 0.5 + 0.5 \times 2.5) \times \frac{1{+}4{\times}50{\times}10^{-2}{\times}5}{4{\times}50{\times}10^{-2}{\times}5} \times \frac{1}{\sqrt{2}} = 13.9V \quad (\textit{The choice: } \boldsymbol{U_S = 15V_{RMS}})$$



The slow rise of the output voltage (turning on) is realized by the circuit around transistor Q601 (the time constant is determined by the resistor R603 and the capacitor C603).

The power dissipation of the IC voltage regulator is: $P_d = (U_{in} - U_{out}) \times I_{Load}$.

$U_{in} - U_{out} > 3V$ *and* $I_{Load} = 1.5A$: $P_d > 3V \times 1.5A = 4.5W$; $I_{Load} = 2.5A$: $P_d > 3V \times 2.5A = 7.5W$.

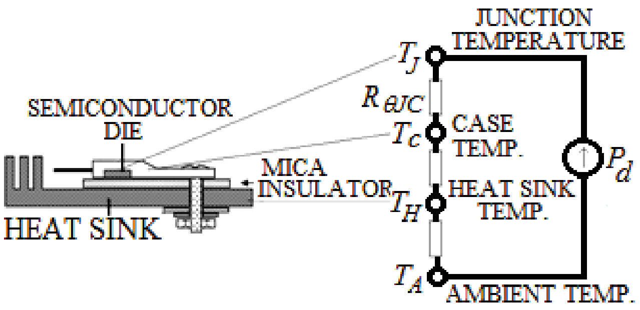
Thermal characteristics of LM 317 and LM338 (TO220 package):

$R_{\theta\,JA}$ *- Junction to ambient thermal resistance without heat sink: 50* $^0C / W$
$R_{\theta\,JC}$ *- Junction to case thermal resistance: 5* $^0C / W$
T_{MAX} *- Maximum junction temperature*

Without any heat sink, the power dissipation of the LM317 is:
(safe operation – junction temperature $T_J = 85\ ^0C$ *and ambient temperature* $T_A = 25\ ^0C$*)*

$$P_d = \frac{(T_J - T_A)}{R_{\theta\,JA}} = \frac{(85-25)C^0}{50C^0/W} = 1.2W$$

Since the power dissipation of the LM317 in the example above is higher than 4.5W (7.5W), it is necessary to mount it on the heat sink.



The thermal circuit of an IC mounted on the heat sink is - thermal circuit equation:

$$\frac{(T_J - T_A)}{P_d} = R_{\theta\,JC} + R_{\theta\,CH} + R_{\theta\,HA}$$

$R_{\theta\,JC}$ *- Junction to case thermal resistance*
$R_{\theta\,CH}$ *- Case to heat sink thermal resistance*
(if mica insulator and thermal grease are used:
$R_{\theta\,CH} = 1.2\ ^0C / W$*)*
$R_{\theta\,HA}$ *– Heat sink to ambient thermal resistance*

The thermal resistance of the heat sink to the ambient equals:

$$R_{\theta\,HA} \le \frac{(T_J - T_A)}{P_d} - R_{\theta\,JC} - R_{\theta\,CH}$$

In the above example ($T_J = 90\ ^0C$ *– safe operation;* $T_A = 25\ ^0C$):

$$I_{Load} = 1.5A: \quad \mathbf{R_{\theta\,HA}} \le \frac{(T_J - T_A)}{P_d} - R_{\theta\,JC} - R_{\theta\,CH} = \frac{90-25}{4.5} - 5 - 1.2 = \mathbf{8.2\ ^0C/W}$$

For safety, choose 5 ^{0}C/W heat sink:

$$I_{Load} = 2.5A: \quad \mathbf{R_{\theta\ HA}} \leq \frac{(T_J - T_A)}{P_d} - R_{\theta\ JC} - R_{\theta\ CH} = \frac{90-25}{7.5} - 5 - 1.2 = \frac{\mathbf{2.4\ ^0C}}{\mathbf{W}}; \quad A\ 2.5\ ^0C/W$$ heatsink is chosen.

In practice, a simplified equation can be used:

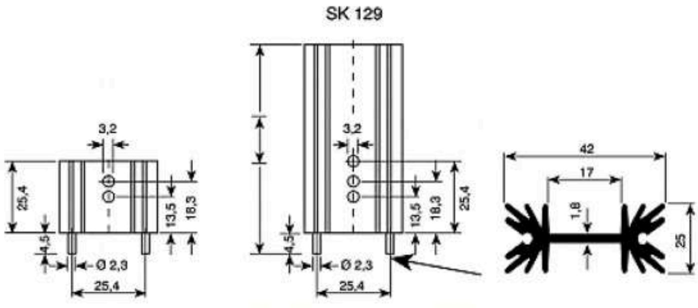
$$R_{\theta\ HA} \leq \frac{(60^0 - T_A)}{P_d}$$

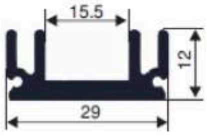
Temperature check:
$I_{Load} = 1.5A$
$T_A = 25\ ^0C$
$T_H = T_A + R_{\theta\ HA} \times P_d = 25\ ^0C + 5\ ^0C / W \times 4.5W = 47.5\ ^0C$
$T_J = T_H + (R_{\theta\ HC} + R_{\theta\ JC}) \times P_d = 47.5\ ^0C + (1.2\ ^0C / W + 5\ ^0C /W) \times 4.5W = 75.4\ ^0C$

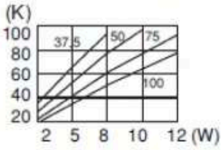
$I_{Load} = 2.5A$
$T_A = 25\ 0C$
$T_H = T_A + R_{\theta\ HA} \times P_d = 25\ ^0C + 2.5\ ^0C / W \times 7.5W = 43.75\ ^0C$
$T_j = T_H + (R_{\theta\ HC} + R_{\theta\ JC}) \times P_d = 43.75\ ^0C + (1.2\ ^0C /W + 5\ ^0C / W) \times 7.5\ W = 90.25\ ^0C$

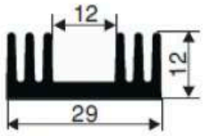
Mounting the IC (TO220 package) on the heatsink
Adequate heat dissipation of the voltage regulator (TO220 package) can be realized by mounting it on the heatsink using parts that make good mechanical and thermal contact between the voltage regulator (TO220 package) and the heat sink.

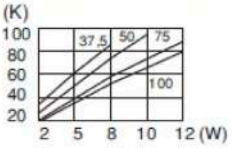
Examples of heat sinks that can be found on the market

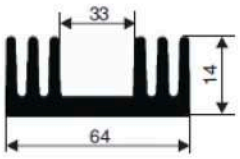


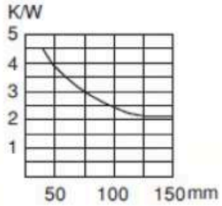


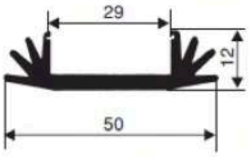


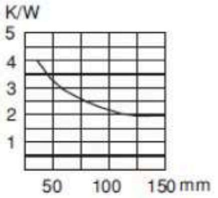


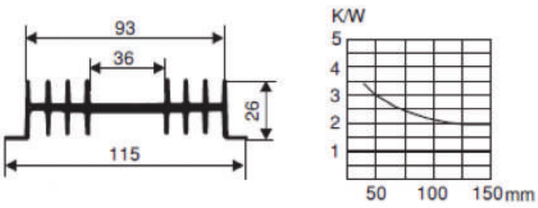


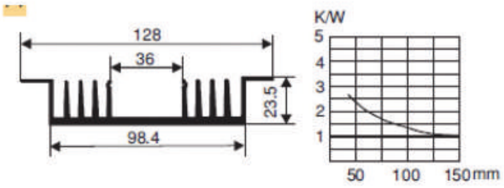


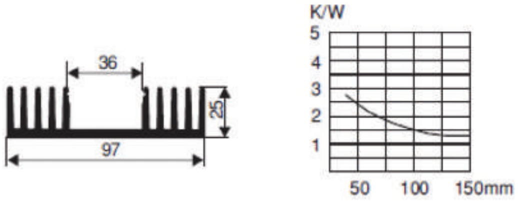


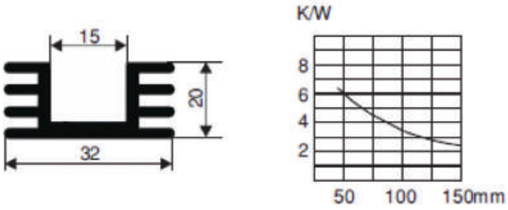


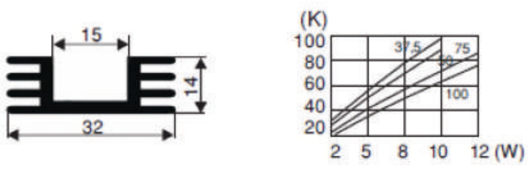


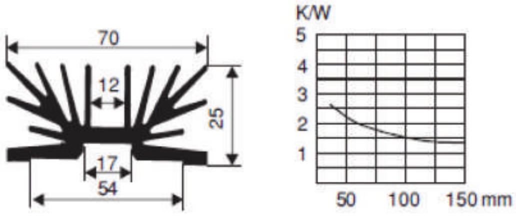


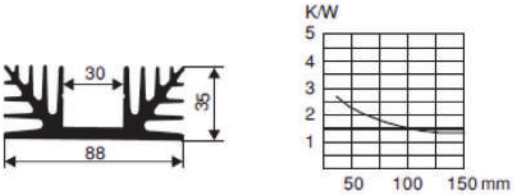


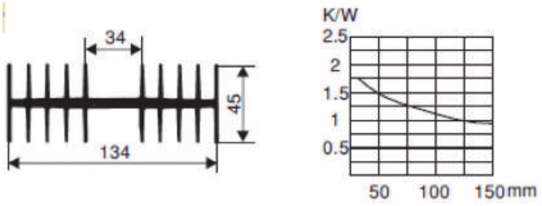




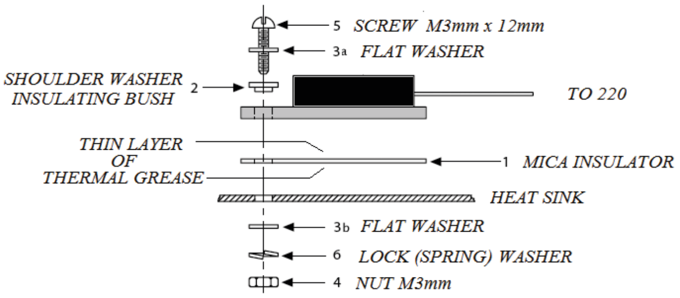


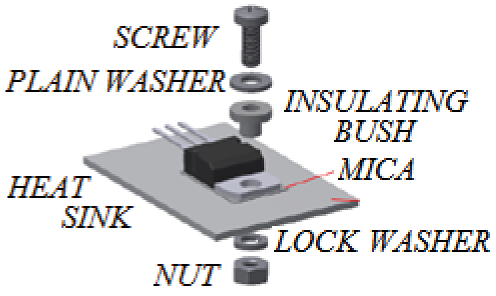






Example of mounting the IC (TO220 package) on the heat sink:





List of electronic components:

ww

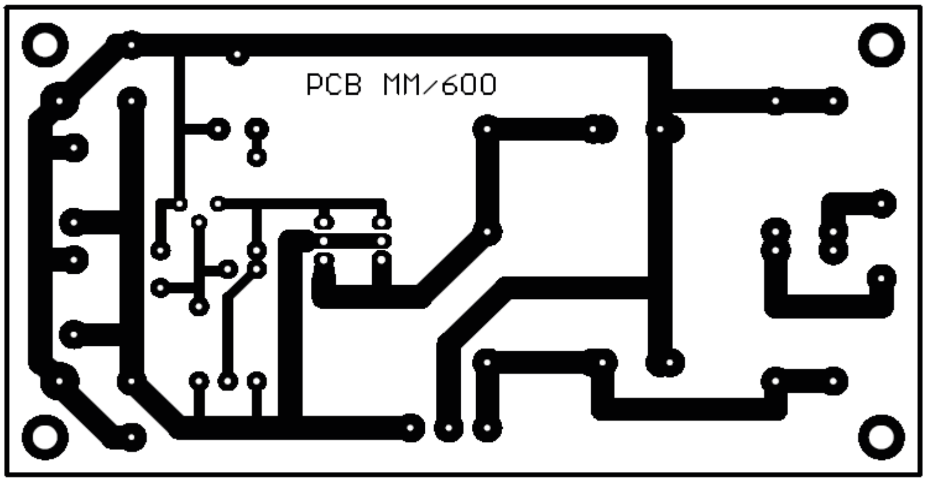
	6.3V 1.5A	12.6V 1.5A	6.3V 2.5A	12.6V 2.5A	5V 1.3A
D601; D602; D603; D604	1N5822	1N5822	1N5822	1N5822	1N5822
D605	1N4002	1N4002	1N4002	1N4002	1N4002
C601; D602;	10.000 µF 16V	10.000 µF 25V	10.000 µF 16V	10.000 µF 25V	10.000 µF 16V
C603	100 µF 16V	100 µF 16V	100 µF 16V	100 µF 16V	100 µF 16V
C604	4.700 µF 16V	4.700 µF 16V	4.700 µF 16V	4.700 µF 16V	4.700 µF 16V
C605	4.700 µF 16V	4.700 µF 16V	4.700 µF 16V	4.700 µF 16V	4.700 µF 16V
R600	1Ω / 10W	1Ω / 10W	0.5Ω / 10W	0.5Ω / 10W	1Ω / 10W
R601	680Ω	1k8	270Ω	820Ω	270Ω
P601	470Ω Trim. Pot.	470Ω Trim. Pot.	470Ω Trim. Pot.	470Ω Trim. Pot.	200Ω Trim. Pot.
R602	220Ω	220Ω	120Ω	120Ω	120Ω
R603	51kΩ	51kΩ	51kΩ	51kΩ	51kΩ
Q601	BC557B 2N2907	BC557B 2N2907	BC557B 2N2907	BC557B 2N2907	BC557B 2N2907
IC601	LM317	LM317	LM338	LM338	LM338
HS601	5^0C / W	5^0C / W	(2.5 … 3.5) ^{0}C / W (5^0C / W for 1.6A)	(2.5 … 3.5) ^{0}C / W (5^0C / W for 1.6A)	5^0C / W
PCB600					
$*U_S$	$10V_{RMS}$	$15V_{RMS}$	$10V_{RMS}$	$15V_{RMS}$	$9V_{RMS}$

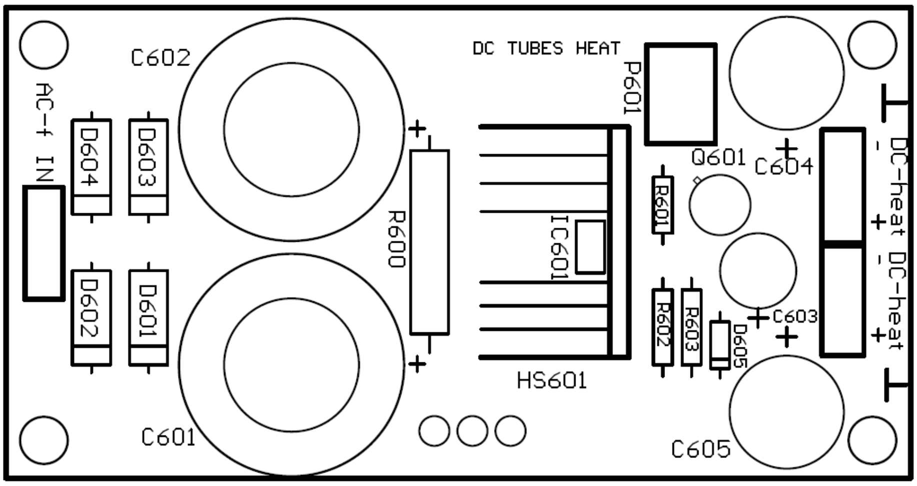
Characteristics of the regulated DC power supply described in the example above: adjustable output voltage, low output hum and noise and low output impedance. It can be used as a DC heater power supply, both for tubes with indirectly heated cathode and for directly heated tubes:

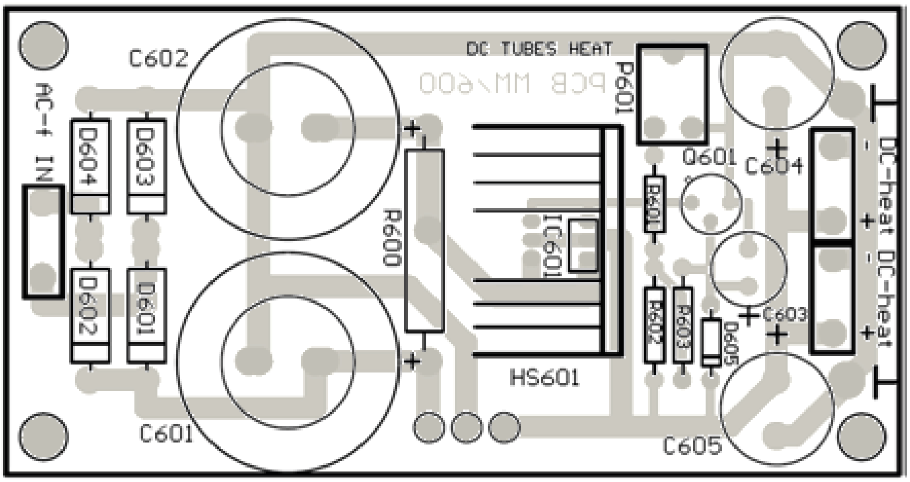
6.3 V (12.6 V)/ 1.5A: Four (Eight) ECC8x, Two (Four) 6SN7, … one (two) EL84…

6.3 V (12.6 V)/ 2.5 A: EL34, KT88, 6L6… (Two: EL34, KT88, 6L6…)

5 V / 1.3 A: 300B


PCB MM/600


C602
AC-f IN
D604
D603
D602
D601
R600
+
+
DC TUBES HEAT
P601
Q601
C604
R601
IC601
DC-heat
DC-heat
C603
R602
R603
D605
HS601
C601
C605


C602
DC TUBES HEAT
PCB MM/600
AC-f IN
D604
D603
D602
D601
R600
P601
Q601
C604
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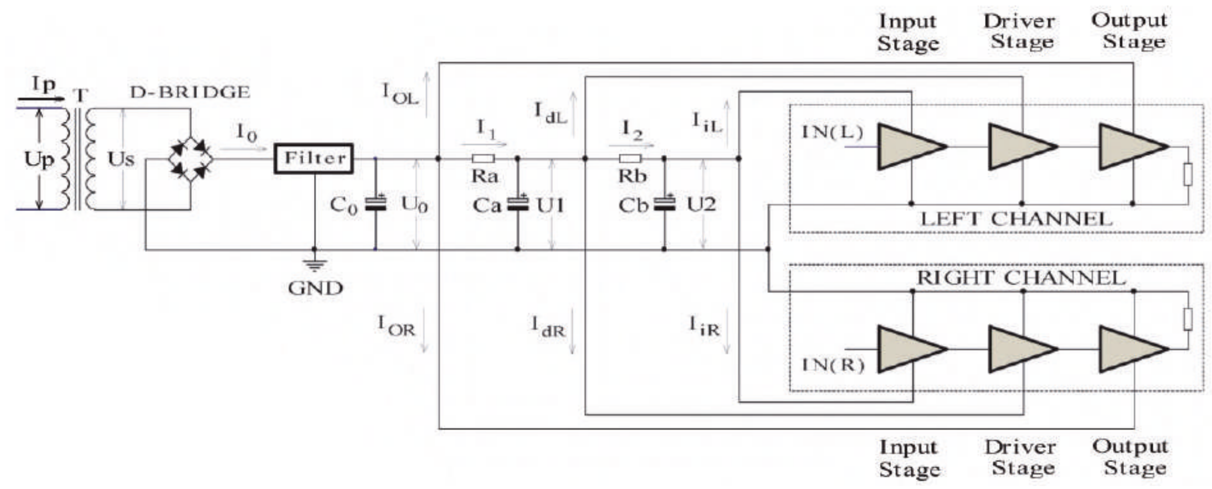
IC601
DC-heat
DC-heat
C603
R602
R603
D605
HS601
C601
C605

Examples of a high-voltage power supply



High-voltage power supply concept:
A common high voltage power supply is used to supply the output stage, driver stage and input stage.

$$I_0 = (I_{OL} + I_{OR}) + I_1 + I_2 = (I_{OL} + I_{OR}) + (I_{dL} + I_{dR}) + (I_{iL} + I_{iR})$$

$$U_1 = U_0 - R_a \times (I_1 + I_2) = U_0 - R_a \times [(I_{dL} + I_{dR}) + (I_{iL} + I_{iR})]$$

$$U_2 = U_1 - R_b \times I_2 = U_1 - R_b \times (I_{iL} + I_{iR})$$

The high voltage power supply of the output stage of the push-pull amplifier does not require a high level of filtering. A simple full wave or bridge rectifier with a filter capacitor can be used.

Example:
Push Pull amplifier
Output stage (one channel), high voltage supply: $U_0 = 420V$, $I_O = 2 \times 90mA$
Driver stage (one channel), high voltage supply: $U_1 = 360V$, $I_d = 10mA$
Input stage (one channel), high voltage supply: $U_2 = 260V$, $I_i = 3.5mA$

Calculation:

$$I_0 = (I_{OL} + I_{OR}) + I_1 + I_2 = (I_{OL} + I_{OR}) + (I_{dL} + I_{dR}) + (I_{iL} + I_{iR})$$

$$I_0 = 2 \text{ x } [\,(2 \text{ x } 90) + 10 + 3.5\,] = 2 \text{ x } 193.5mA = 387mA \approx 400mA$$

$$R_1 = R_a = \frac{U_0 - U_1}{I_1 + I_2} = \frac{U_0 - U_1}{2 \times I_d + 2 \times I_i} = \frac{420 - 360}{2 \times 10 + 2 \times 3.5} = 2.2k\Omega$$

$$P_{d\,(R1)} = R_1 \text{ x } (I_1 + I_2)^2 = 2200 \text{ x } 0.027^2 = 1.6W \quad (R_1 = 2.2k\Omega\,/\,7W)$$

$$R_2 = R_b = \frac{U_1 - U_2}{I_2} = \frac{U_1 - U_2}{2 \times I_i} = \frac{360 - 260}{2 \times 3.5} = 14.28k\Omega$$

Checking the ripple factor (RF):

Filter capacitors:
C_o *(two capacitors connected in series):* $C_1 = C_1' = 2200\mu F / 350V \rightarrow C_{1\,eq} = C_1 / 2 = 2200 / 2 = 1100\ \mu F$
Ca : $C_2 = 470\mu F / 400V$
Cb : $C_3 = 220\mu F / 400V$

Ripple factor (U_0 = 420V and I_0 = 400mA):

$$\mathrm{f} = 50\ \mathrm{Hz}: \quad RF = \frac{U_{r\,(RMS)}}{U_{DC}} = \frac{1}{4\times\sqrt{3}\times f\times C\times R_L} = \frac{2.88}{C\times R_L} = \frac{2.88}{C\times\frac{U_0}{I_0}} = \frac{2.88}{1100\times\frac{420}{400}} = \mathbf{0.00249} \quad \mathrm{C}\ [\mu\mathrm{F}],\ \mathrm{R}\ [\mathrm{k}\Omega]$$

Ripple voltage (RMS) is:

$$U_{r\,(RMS)} = RF \times U_{DC} = 0.00249 \times 420 = 1V$$

Ripple voltage (U_1 = 360V and I = 27mA):

To simplify the calculation of the ripple voltage of U_1, an equivalent circuit consisting of an AC voltage source of $1V_{RMS}$ (ripple voltage of U_0) and a voltage divider R_1 and X_{C2} can be used:

The impedance of capacitor C_2 is: $X_{C2} = 1 / (2\pi fC_2) = 1 / (2x3.14x100x470x10^{-6}) = 3.388\Omega$ (f = 100Hz)

$$U_{r\,(RMS)} = 1V \times \frac{X_{C2}}{X_{C2} + R_1} = 1V \times \frac{3.388}{3.388 + 2200} = 1.5mV$$

Ripple voltage (U_2 = 260V and I = 7mA):
Repeating the previous procedure: $1.5mV_{RMS}$ AC voltage source (ripple voltage of U_1) and a voltage divider R_2 and X_{C3} .

The impedance of capacitor C_2 is: $X_{C2} = 1 / (2\pi fC_2) = 1 / (2x3.14x100x220x10^{-6}) = 7.24\Omega$ (f = 100Hz)

$$U_{r\,(RMS)} = 1.5mV \times \frac{X_{C3}}{X_{C3} + R_2} = 1V \times \frac{7.24}{7.24 + 12000} = 0.9\mu V$$

Calculation of the secondary winding voltage of the power transformer:

If a long life (LL) grade aluminum electrolytic capacitor is used:

$$\boldsymbol{U_{S\,[RMS]}} = \frac{\boldsymbol{U_{out} + I_{Load}\times(R_{S[DC]} + n^2\times R_{P[DC]}) + 2\times U_D}}{\sqrt{2}}, \quad R_P \approx 5\Omega,\ n \approx 1.45,\ R_S = 15\Omega,\ U_D \approx 0.8V$$

$$U_{S\,[RMS]} = \frac{420 + 0.4 \times (15 + 1.45^2 \times 5) + 2 \times 0.8}{\sqrt{2}} = 306V \approx 310V$$

The Secondary winding must withstand a current of: $I_{S\,[RMS]} \approx 1.5 \times I_{DC\,[Load]}$ *, ($I_{S\,[RMS]} \approx 1.5$ x 400mA = 600mA)*

Diodes
V_{RRM} (Repetitive Reverse Voltage), bridge rectifier:

$$V_{RRM} > \sqrt{2}\times U_S = 1.41\times U_S = \sqrt{2}\times 310 = 452.5$$ *(choice: Si diode, V_{RRM} = 1000V)*

$I_{F(AV)}$ (Average rectified forward current):
$I_{F(AV)} > I_{Load}$ = 400mA, (choice: Si diode, $I_{F(AV)}$ = 3A),
[reminder: if the diode is mounted on the PCB – the safe operating current of the diode is about half the $I_{F(AV)}$ i.e. 1.5A].
Fast Si rectifier diodes that fulfill the above requirements: BYT 56M, UF5408, (1N5408).

Balancing resistors *(connected in parallel with input capacitors):*

$$I_{leak}[\mu A] = 2.5\times 10^{-4}\times C[\mu F]\times U_R[V] + 1\,\mu A = 2.5\times 10^{-4}\times 2200\times 350 + 1 = 193.5\,\mu A$$

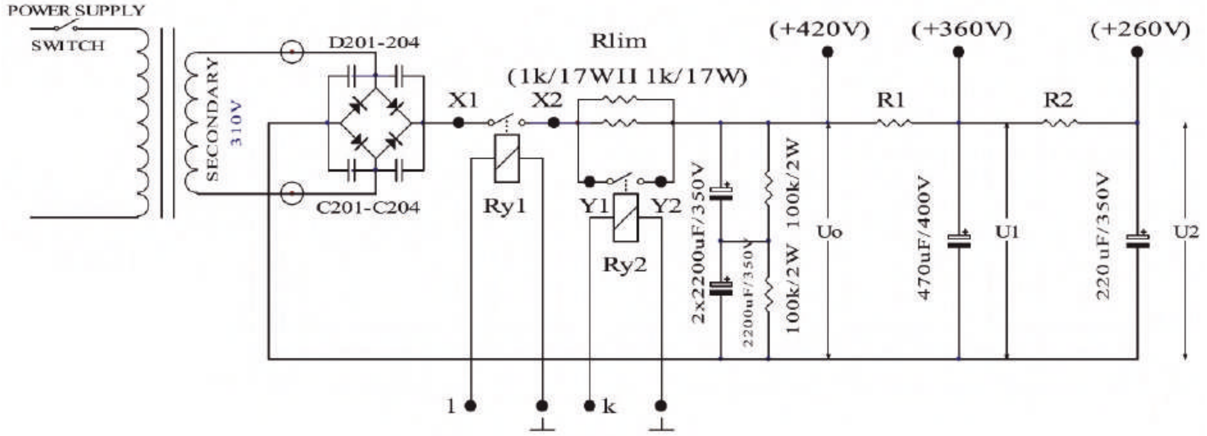
$I_{Bal} = 20 \times I_{leak}$ = 20 x 193.5 = 3870μA = 3.87mA
2 x R_{Bal} = (applied voltage) / I_{Bal} = 420V / 3.87mA = 108kΩ → R_{Bal} = 54kΩ (choice: 100kΩ / 2W)

Delay circuit*: the same circuit is used as the circuit detailed in the text above (Practical delay circuit).*

***Capacitors connected in parallel with the rectifier bridge diodes**: 100nF / 630V*

Secondary winding of power transformer:

U_S =310V, I_S = 600mA, $P_S = U_S \times I_S$ = 310V x 0.6A = 186W



Single ended amplifiers require very high quality filtering of high voltage. In practice (usually) a simple CLC input filter can be applied.

Example:
Single Ended amplifier
Output stage (one channel), high voltage supply: U_0 = 430V, I_O = 81mA
Driver stage and input stage (one channel), high voltage supply: U_1 = 380V, I_d = 10mA

Calculation:
$I_0 = (I_{OL} + I_{OR}) + I_1 = (I_{OL} + I_{OR}) + (I_{dL} + I_{dR})$
$I_0 = 2 \times [81 + 10] = 2 \times 91mA = 182mA \approx 200mA$
R_{Load} = 430V / 200mA = 2150Ω

$$R_1 = R_a = \frac{U_0 - U_1}{I_1 + I_2} = \frac{U_0 - U_1}{2 \times I_d + 2 \times I_t} = \frac{430 - 380}{2 \times 10} = 2.5k\Omega$$

$P_{d\ (R1)} = R_1 \times I_1^2 = 2500 \times 0.02^2 = 1W$, (R_1 – 2.4kΩ / 5W)

Checking the ripple factor (RF)
CLC filter:
C_1 = 2 x 470 **µF / 350** V connected in series → $C_{1\ eq} = C_1 / 2 = 470 / 2 = 235$ µF
C_2 = 2 x 470 **µF /400** V connected in series and connected in parallel with 2 x 470 **µF /400** V connected in series, $C_{2\ eq}$ = 470 **µF**
C_3 = 2 x 330 **µF /400** V connected in series → $C_{3\ eq}$ = 165 µF
L = 5H / 250mA / $R_{choke\ -\ DC}$ = 55 Ω

Ripple factor (U_0 = 430 V and I_0 = 200 mA *):*

$$RF = \frac{\sqrt{2}}{8 \times (2 \times \pi \times f)^3 \times L \times C_1 \times C_2 \times R_{Load}} = \frac{\sqrt{2}}{8 \times (2 \times \pi \times 50)^3 \times 5 \times 235 \times 10^{-6} \times 470 \times 10^{-6} \times 2150} = 4.8 \times 10^{-6}$$

Ripple voltage (RMS):

$$U_{r\ (RMS)} = \frac{\sqrt{2} \times U_{S\ peak}}{(4 \times \pi \times f)^3 \times L \times C_1 \times C_2 \times \left[1 + \frac{1}{4 \times f \times C_1 \times R_{Load}}\right] \times (R_{Load} + R_{choke-DC})} = 2\text{mV}, or:$$

$$U_{r\ (RMS)} = RF \times U_{DC} = 4.8 \times 10^{-6} \times 430 = 2mV$$

Ripple voltage (U = 380 V and I = 20 mA):

AC voltage source of 2mV$_{RMS}$ and a voltage divider R_2 and X_{C3} .
Impedance of C_3 capacitor is: $X_{C3} = 1 / (2\pi f C_3) = 1 / (2x3.14x100x165x10^{-6}) = 9.65\ \Omega$ (f = 100Hz)

$$U_{r\,(RMS)} = 2mV \times \frac{X_{C3}}{X_{C3} + R_2} = 2mV \times \frac{20.617}{20.617 + 2400} = 0.004mV$$

Calculation of the secondary winding voltage of the power transformer:

If a long-life (LL) grade aluminum electrolytic capacitor is used:

$$U_{S\,[RMS]} = \frac{U_{out} + I_{Load} \times (R_{S[DC]} + n^2 \times R_{P[DC]}) + 2 \times U_D}{\sqrt{2}}, \quad R_P \approx 5\Omega,\ n \approx 1.45,\ R_S = 15\Omega,\ U_D \approx 0.8V$$

$$U_{S\,[RMS]} = \frac{430 + 0.2 \times (15 + 1.45^2 \times 5) + 2 \times 0.8}{\sqrt{2}} = 309.7V \ \approx 310V$$

The Secondary winding must withstand a current of: $I_{S\,[RMS]} \approx 1.5 \times I_{DC\,[Load]}$ *, ($I_{S\,[RMS]} \approx 1.5 \times 200$ mA = 300 mA)*

Diodes
V_{RRM} (Repetitive Reverse Voltage), bridge rectifier:
$V_{RRM} > \sqrt{2} \times U_S = 1.41 \times U_S = \sqrt{2} \times 310 = 452.5V$ *(choice: Si diode, V_{RRM} = 1000V)*
$I_{F(AV)}$ (Average rectified forward current):
$I_{F(AV)} > I_{Load}$ = 200mA, choice: Si diode, $I_{F(AV)}$ = 3A,
[reminder: if the diode is mounted on the PCB – the safe operating current of the diode is about half the $I_{F(AV)}$ i.e. 1.5A].
Fast Si rectifier diodes that fulfill the above requirements: BYT 56M, UF5408, (1N5408).

Balancing resistors *(connected in parallel with input capacitors):*

$$I_{leak}[\mu A] = 2.5 \times 10^{-4} \times C[\mu F] \times U_R[V] + 1\,\mu A = 2.5 \times 10^{-4} \times 470 \times 350 + 1 = 42\,\mu A$$

470 μF connected in series with 470 μF

$I_{Bal} = 20 \times I_{leak} = 20 \times 42 = 842$ μA = 0.842 mA
2 x R_{Bal} = (applied voltage) / I_{Bal} = 430 V / 0.842 mA = 510 kΩ → R_{Bal} = 255 kΩ , (choice: 270 kΩ / 2W)

Second group of capacitors (470 μF connected in series with 47 0μF and connected parallel with 470 μF connected in series with 47 0μF).

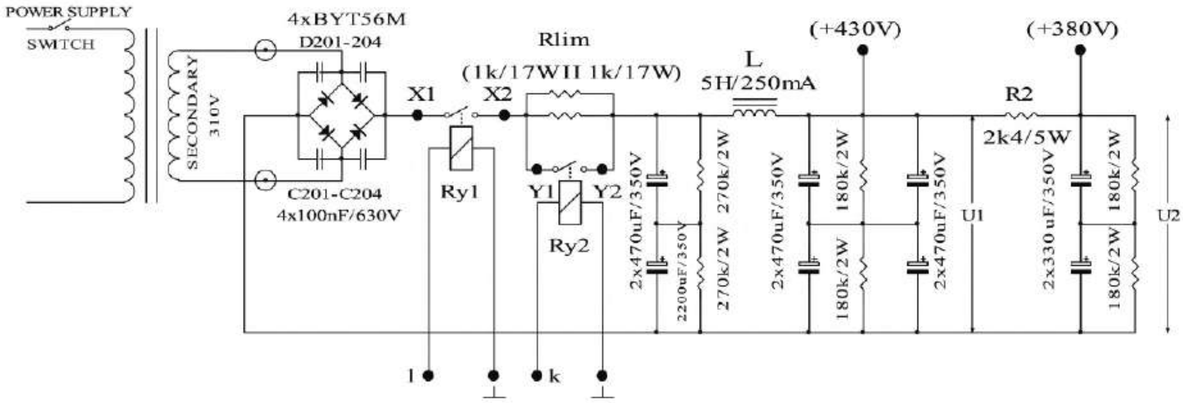
$R_{Bal} = \frac{100}{C[\mu F] \times \sqrt{n}}\ [M\Omega] = \frac{100}{470 \times \sqrt{2}} = 150.9\ k\Omega$, *(choice is: 180 kΩ / 2W)*

Delay circuit*: the same circuit is used as the circuit detailed in the text above (Practical delay circuit).*

Capacitors connected in parallel with the rectifier bridge diodes*: 100 nF / 630 V*

Secondary winding of power transformer:

U_S =310 V, I_S = 300 mA, $P_S = U_S \times I_S$ = 310 V x 0.3 A = 93 W



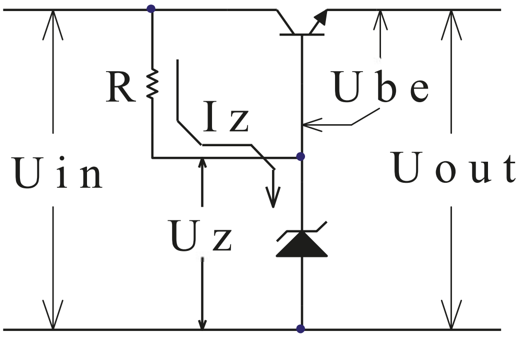
6.7 REGULATED HIGH-VOLTAGE POWER SUPPLIES

A regulated high voltage power supply is used to power the preamplifiers, the input and drive stage of the amplifiers or the screen grid (g2) of the tetrode or pentode of the amplifier output stage, usually (can be used to power the output stage of the amplifier, but the design requirements such as high voltage and high current complicate its design).

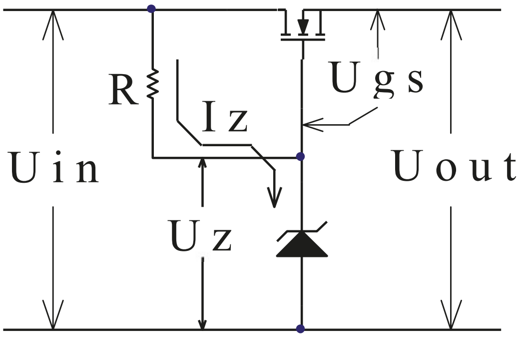
The series voltage regulator is a simple and very effective regulated high voltage power supply.

Basic configuration:

Emitter follower regulator



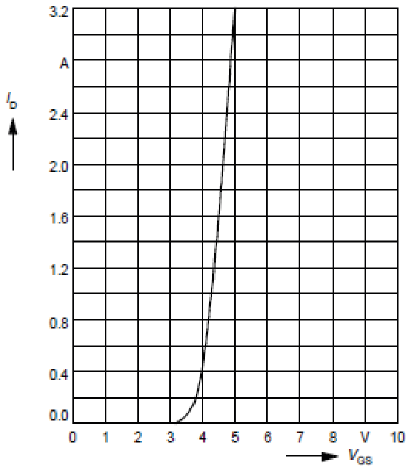
Source follower regulator



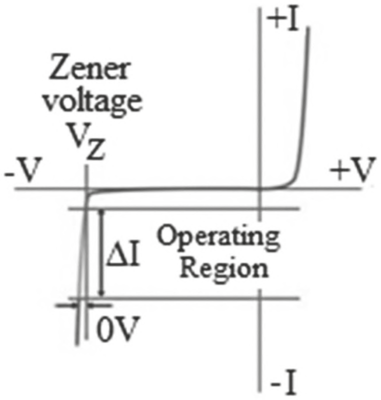
The principle of operation is very simple. A high-voltage, high-power active electronic component such as a transistor (BJT) or MOSFET is connected in series to the load. The base or drain is kept at a constant voltage by a Zener diode (constant voltage: U_Z). If the output voltage decreases, the base to emitter voltage increases and the current flowing through the transistor and the load also increases, and the output voltage increases i.e. the decrease of the output voltage is compensated. A similar, but opposite, process occurs when the output voltage is increased.

The basic equations:

$$U_{out} = U_z - U_{be} \quad \text{or,} \quad U_{out} = U_z - U_{gs} \qquad I_z = \frac{U_{in} - U_z}{R} - I_b$$



MOSFET, drain current vs gate to Source voltage.



Zener diode characteristics

U_Z - rated voltage of Zener diode

U_{be} - base to emitter voltage
(Si transistor: U_{be} is about 0.7V)

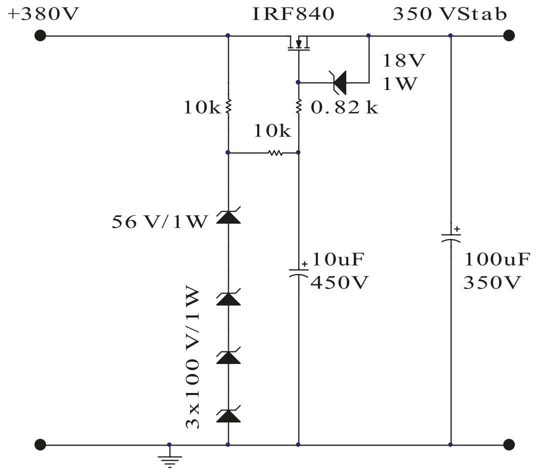
U_{gs} - gate to source voltage
(it is necessary to consult the MOSFET data sheet, usually about 3.5V)

I_b – transistor base current

$$I_b = \frac{I_E}{(1+h_{fe})}$$

** h_{FE} of a transistor is the current gain or amplification factor of a transistor*

Example

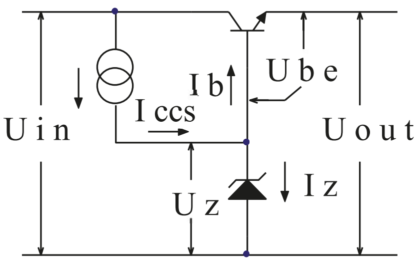


Simple series voltage regulator using high-voltage, high-power MOSFET: U_{out} = 350V, I_{out} = 200mA
High voltage N channel MOSFET: IRF 840 500V, 8A
$U_{GS} \approx 3.5V$ (I_D = 200mA)
$U_{out} = U_Z - U_{gs}$ → $U_Z = U_{out} + U_{gs} = 350 + 3.5 = 353.5V$

Combination of Zener diodes connected in series:
3 x 100V + 1 x 56V = 356V
Expected output voltage: U_{out} = 356V - 3.5V = 352.5V
1W Zener diode:
56V: $I_{Z\ Test}$ = 4.5mA; 100V: $I_{Z\ Test}$ = 2.5mA
U_{in} = 380V. The choice of I_Z is around 3mA.

$$R = \frac{U_{in}-U_Z}{I_Z} = \frac{380-356}{3} = 8k\ ;\ (choice: R = 10k)$$

Variations of the input unregulated voltage and variations of the base current of the transistor, even if small, cause a change of the current flowing through the Zener diode and changes of the Zener voltage (as can be seen in the Zener diode characteristics diagram).



In practice, this problem of a simple series voltage regulator is solved by using a constant current source to supply Zener diode ($I_{CCS} = I_Z + I_b$). High voltage power transistors usually have a low gain (h_{fe} around 30) and the base current is not negligible (it can be several mA) so the power dissipation of the constant current source is not negligible. As the gate current of the high voltage MOSFET is in the order of nA, its application in practice simplifies the design of a CCS and is used more often.

The U_{CE} voltage of the high voltage power transistor must be higher than the unregulated voltage

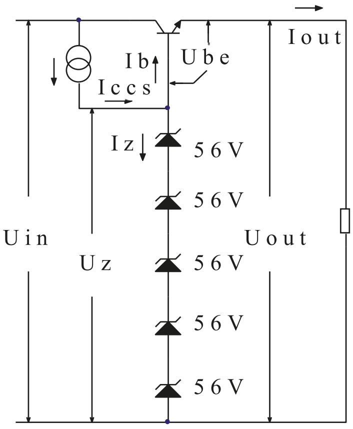
$U_{unregulated} - U_{CE} > U_{unregulated}$ and the collector current must be higher than the load current - $I_C > I_{Load}$ (MOSFET: $U_{DS} > U_{unregulated}$ and $I_D > I_{Load}$). Examples of high voltage power transistor and MOSFET that can be used in a high voltage series voltage regulator circuit:

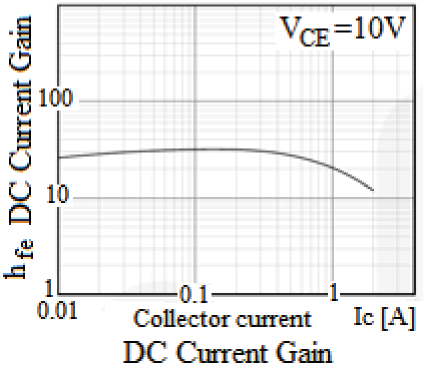
Transistor TIP 50: $U_{CE} = 400V$ *DC,* $I_C = 1A$, $h_{fe} \approx 30$
MOSFET (SIPMOS) BUZ 92: $U_{DS} = 600V$, $I_D = 3.3A$

Zener diode*: If the designed voltage and power of the Zener diode cannot be realized using one standard Zener diode, several standard Zener diodes can be connected in series with a total voltage equal to the designed and a total power higher than the designed.*

A high voltage transistor must be used in the constant current source circuit:
$U_{CE} > U_{unregulated}$ *and* $I_C > I_Z + I_b$ *(MOSFET series regulator: collector current of current source transistor* $I_C > I_Z$*).*

Example:
Series voltage regulator: $U_{out} = 280V$, $I_{out} = 100\ mA$.
Transistor: TIP 50, $U_{CE} = 400V$, $I_C = 1A$, $P_C = 40W$





DC Current Gain

From the diagram:
$I_C = 0.1A$, $h_{fe} = 30$
Basic equations of the transistor:
$I_C = h_{fe} \times I_b$ *and* $I_E = (h_{fe} + 1) \times I_b$,

Transistor base current is:

$$I_b = \frac{I_e}{h_{fe} + 1}$$

$$I_b = \frac{100mA}{30 + 1} - 3.2mA$$

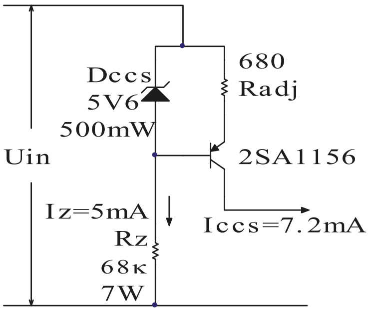
Zener diode: 5 x Zener diodes of $V_Z = 56V$ *connected in series, total* $V_Z = 5 \times 56V = 280V$.
1.3W Zener diode BZX85C56, $I_{Z\ TEST} = 4mA$, *for* $I_Z = 4mA$ *the dissipation of the diode is 56V x 0.004A = 0.224W*
Some other combination of Zener diodes connected in series can be used.

Constant current source (CCS):

The current of the constant current source is: $I_{CCS} = I_Z + I_b = 4 + 3.2 = 7.2mA$

Transistor: PNP 2SA1156, $U_{CE} = 400V$, $I_C = 500mA$, $P_C = 10W$

Simple CCS circuit:



CCS current: $I_{CCS} = \frac{V_{Z\,(Dccs)} - U_{be}}{R_{adj}} \rightarrow R_{adj} = \frac{V_{Z\,(Dccs)} - U_{be}}{I_{CCS}}$

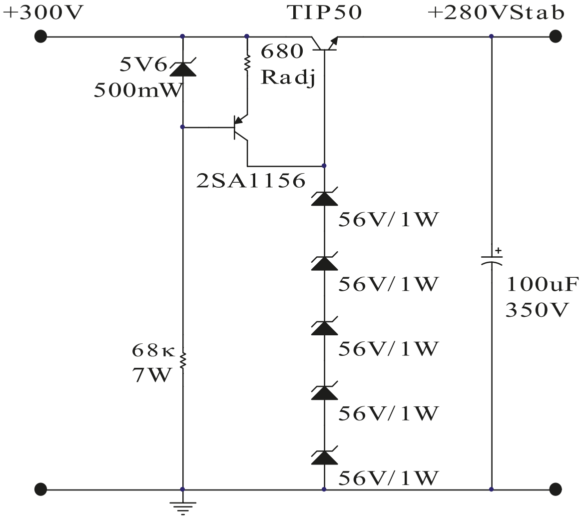
5.6V / 500mW Zener diode: BZX79V5V6

$$R_{adj} = \frac{5.6-0.7}{7.2} = 0.68k\Omega$$

For $U_{in} = 300V$ *and* $I_Z = (3 \ldots 5)mA$ *of the Zener diode BZX79V5V6:*

$R_Z = \frac{U_{in} - V_{Z\,(Dccs)}}{I_Z} = \frac{300-5.6}{5} = 59k\Omega$, (choice: $R_Z = 68k\Omega / 7W$).

The synthesized schematic of the regulated voltage 280V / 100mA power supply is:



Power dissipation of the transistor TIP 50:

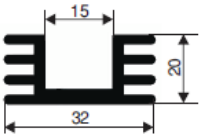
$P_d = (U_{in} - U_{out}) \times I_{Load} = (300 - 280) \times 0.1 = 2W$

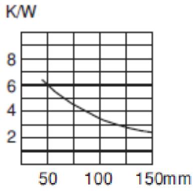
Taking into account the tolerance of the heat sink characteristic, the heat sink is calculated for $P_d = 5W$*:*

$$R_{\theta\,HA} \le \frac{(T_J - T_A)}{P_d} - R_{\theta\,JC} - R_{\theta\,CH} = \frac{90-25}{4.5} - 3.125 - 1.2 = 8.7^0 C/W$$

The TIP 50 must be mounted on the heat sink:

$R_{\theta\,HA} \le 8^0 C/W$, *30mm long. Example of a heat sink:*





The MOSFET version of the above regulated voltage power supply:

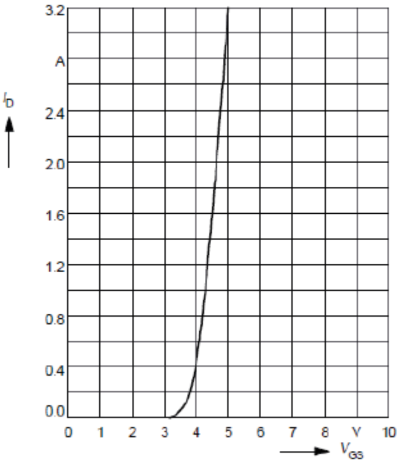
N channel high voltage power MOSFET: BUZ 92, IRF 840, BUK 456 800A, STW12NK95Z …

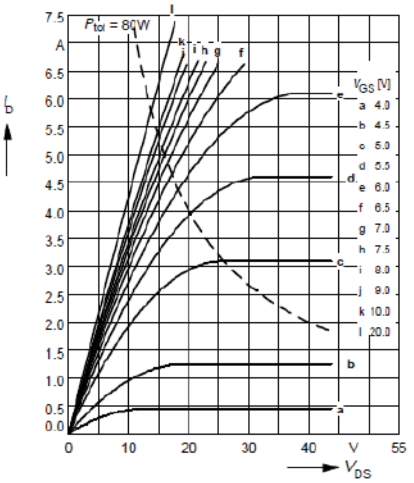
The only difference compared to the BJT regulator is that the CCS is designed to supply only Zener diodes because the MOSFET gate current (around 100nA) is much lower than the base current of the transistor (3.2mA). So, the CCS current is around 4mA. Also, from the MOSFET data sheet and characteristics:

Characteristics $I_D - f(U_{GS})$*: for* $I_D = 100mA \rightarrow U_{GS}$ *is around 3.4V.*

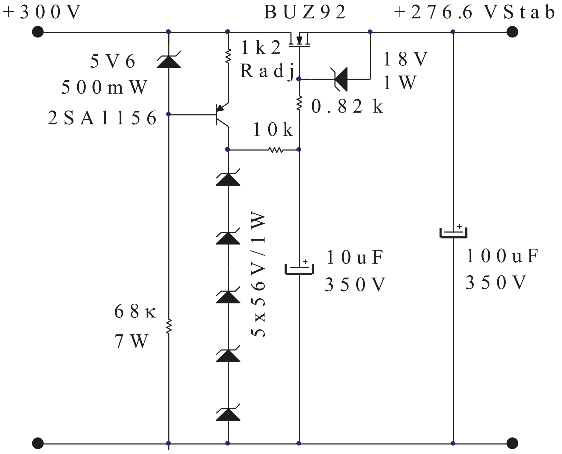
Characteristics $I_D = f(U_{DS})$, $I_D|_{UGS\,=\,Ctc}$*: for* $I_D = 100mA \rightarrow$ *minimum* $U_{DS} \ge 10V$.

BUZ 92: $U_{DS} = 600V$, $I_D = 3.3A$, $P_d = 80W$





CCS: $R_{adj} = (U_Z - U_{bE}) / I = (5.6 - 0.7)V / 4mA = 1k2$

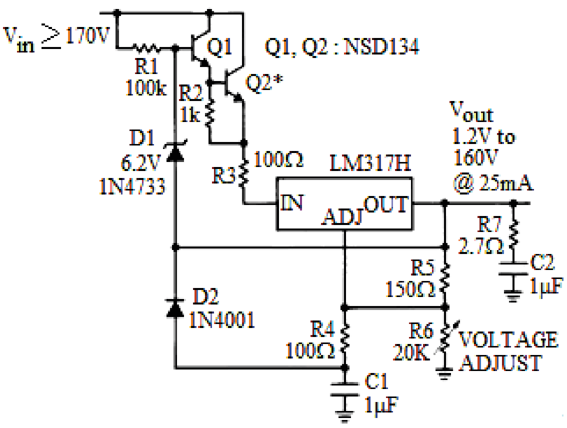


An important characteristic of a series voltage regulator is:

The suppression of AC signals present at the input of the regulator (ripple voltage and its harmonics, hum and noise) is very effective and high (examples above: it is about 46 dB).

6.7.1 HIGH-VOLTAGE ADJUSTABLE POWER SUPPLY - MAIDA TYPE

Published in 1980 by Texas Instruments (LB-47 High Voltage Adjustable Power Supplies)
Author Michael Maida.



The basis of the Maida high voltage regulator is an unusual application of the low voltage IC voltage regulator LM317H:

“The regulator has no ground pin; instead, all quiescent current (about 5 mA) flows to the output terminal. Since the regulator sees only the input - output differential, its voltage rating – 40 V … will not be exceeded for outputs of hundred volts… Zener diode D1 ensures that the LM317 sees only a 5V input-output differential over the entire range of output voltage from 1.2 V to 160 V. Since high voltage – transistors by necessity have a low β, a Darlington is used to stand off the high voltage … R3 limits short circuit current to 50 mA. The RC network on the output improves transient response as does bypassing the ADJUST pin, while R4 and D2 protect the ADJUST pin during shorts…

These techniques can be extended for higher output voltages and/or currents by either using better high voltage transistors or cascading or paralleling (with appropriate emitter ballasting resistors) several transistors.”

Design procedure

1. Design input data: output voltage, current and limit current of high voltage power supply U_{out}, I_{out}, $I_{out\ Limit}$.
   The input data determines the choice of the high voltage transistor Q2 ($U_{CE} > U_{in}$, β, $U_{bE,}$, P_d)

2. Calculation of the voltage of the Zener diode D1:

   Introductory remarks
   According to the above schematic, the following equations of D1 circuit can be derived:

   $U_{z\ d1} = U_{be} + R_3 \times I_{out} + U_{in\text{-}out\ LM317}$
   $U_{in\text{-}out\ LM317} \times I_{out} = P_{d\ LM317}$ (power dissipation of LM317)
   $U_{in\text{-}out\ LM317\ (min)} \geq 3V$, necessary for normal operation of LM317

   If the power dissipation of LM317 is < 1W, the use of a heat sink is not necessary, for power dissipation > 1W, the LM 317 must be mounted on an adequate heat sink.

   So, for example, if the load current is less than or equal to 100mA,

   $$U_{in-out\ LM317} \leq \frac{P_{d\ LM317}}{I_{out}} = \frac{1W}{0.1A} = 10V$$

   If $I_{out} = 200mA$ *and* $U_{in-out} \leq 10V$ *and the LM317 must be mounted on a heat sink.*

   $P_{d\ LM317} = Ui_{n-o\ LM317} \times I_{out} = 10V \times 0.2A = 2W$, *and the LM 317 must be mounted on a heat sink.*

   Calculation:
   The calculation starts with the equation $U_{Z\,D1} = U_{bE} + R_3 \times I_{out} + U_{in-out\,LM317}$ but without the factor of equation $R_3 \times I_{out}$ and with the chosen $U_{in\text{-}out\ LM317}$ determined as explained above

   (P_d = required power dissipation of LM317, $U_{in-out\,LM317} = \frac{P_{d\,LM317}}{I_{out}}$):

   $U_{Z\,D1} \geq U_{bE} + U_{in-o\quad LM317}$ (if a MOSFET is used, substitute U_{bE} with U_{GS} in the equation).

   Choose a standard Zener diode - $U_{Z\ D1\text{-}\ R}$

   Calculation of R_3: $R_3 = \frac{U_{Z\,D1-R} - U_{Z\,D1}}{I_{out}}$

3. Calculation of R_5 and R_6 :

   Note: The minimum output current of the LM317 required for its regular operation is 5mA.
   Under the condition that the high voltage regulator is loaded with I_{Load} > 5mA, the current *flowing through* R_6 *can be less than 5mA (1mA ÷ 2mA).*

   Calculation of R_6 (taking into account the power dissipation of R_6 whose resistance can be very high) (Note: power dissipation of R_6 less than 1W):

   $$R_6 > \frac{U_{out}^2}{P_{R6\ diss}} \quad \text{and} \quad R_6 > \frac{U_{out}\,[V]}{(1-2)mA}; \quad \text{Choose standard R6}$$

   Calculation of R_5: $U_{out} = 1.25 \times \left(1 + \frac{R_6}{R_5}\right) \rightarrow R_5 = \frac{R_6}{\frac{U_{out}}{1.25} - 1}$

   Note: 120Ω ≤ R_5 ≤ 1000Ω

4. Calculation of unregulated input voltage:

   It is necessary to consult the manufacturer's data sheet of the high voltage power transistor or MOSFET. First, the collector and base currents of the high voltage transistor are calculated:

Collector current is: $I_C = h_{fe} \times I_b$

Emitter current is: $I_E = (h_{fe} + 1) \times I_b$

- By analyzing the schematic, the equation can be derived: $I_E = I_{out} + I_{R6} - I_{Z\,D1}$
- $I_b = \frac{I_E}{h_{fe}+1}$ (h_{fe}, using the transistor data sheet)
- $I_C = h_{fe} \times I_b$
- Considering that I_b and I_C have been previously calculated, the collector to emitter voltage U_{CE} can be determined using the characteristic $I_C = f(U_{CE})$ for I_b = Ctc.
- Unregulated input voltage:

  $$U_{in} \geq U_{CE} + U_{out} + U_{Z\,D1} - U_{bE}$$

5. Calculation of R_1 :

   $$R_1 = \frac{U_{in} - (U_{out} + U_{Z\,D1})}{I_{Z\,D1} + I_b}$$

6. Calculation of the power dissipation of the high voltage transistor and LM317 and calculation of their heat sinks, if necessary.

Example
High voltage regulator - Maida type: $U_{out} = 300V$, $I_{out} = 200mA$
*High voltage transistor **MJE 13005**; 75W, 400V, 4A Si NPN.*
***$P_{d\,LM317}$ = 2W** (LM317 mounted on the heat sink)*

$$U_{in-out\,LM317} = \frac{P_{d\,LM317}}{I_{out}} = \frac{2W}{0.2A} = 10V$$

$$U_{Z\,D1} \geq U_{bE} + U_{in-out\,LM317} = 0.7V + 10V = 10.7V$$

*Zener diode 15V / 1W (**$U_{Z\,D1-R}$ = 15V**)*

Note: I_Z >5mA

$$R_3 = \frac{U_{Z\,D1-R} - U_{Z\,D1}}{I_{out}} = \frac{15-10.7}{0.2} = 21.5\Omega; \quad \boldsymbol{R_3\ standard = 22\ \Omega / 5W}$$

$$R_6 > \frac{(U_{out}-1.25)\ [V]}{(1-2)mA} = \frac{(300-1.25)\mathrm{V}}{(1-2)mA} = (149 \div 298.75)\mathrm{k\Omega}; \quad R_6\ standard = 220\ k\Omega$$

$$P_{R6\,diss} = \frac{(U_{out}-1.25)}{R_6} = \frac{(298.75V)^2}{220k\Omega} = 0.406W; \quad \boldsymbol{R_6 - 220\ k\Omega / 2W}$$

$$R_5 = \frac{R_6}{\frac{U_{out}}{1.25} - 1} = \frac{220k\Omega}{\frac{300V}{1.25} - 1} = 920\Omega; \quad \boldsymbol{R_5\ -\ 470\ \Omega + 1k\Omega\ \ trim\ pot}$$

$$I_{R6} = \frac{(U_{out} - 1.25)\ [V]}{R_6} = \frac{(300 - 1.25)V}{220k\Omega} = 1.358mA$$

$$I_E = I_{out} + I_{R6} - I_{Z\,D1} = 200 + 1.358 - 6 = 195.358mA$$

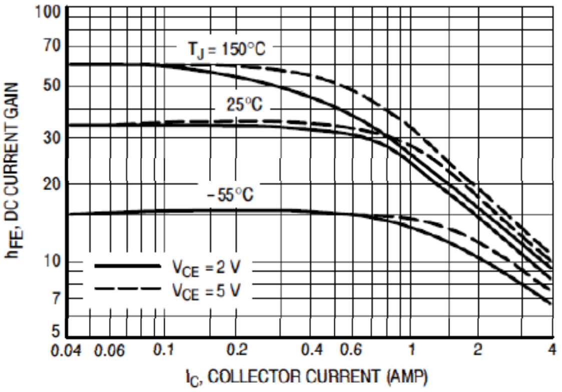


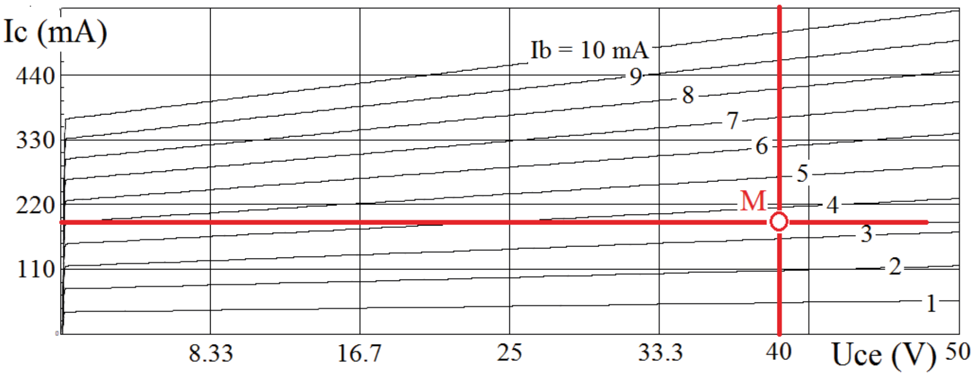
Figure 1. DC Current Gain

$$I_b = \frac{I_E}{h_{fe}+1} = \frac{195.358}{55+1} = \mathbf{3.488mA}$$

Using the diagram $h_{fe} = f(I_C)$: for $I_C = 200mA$, $h_{fe} \approx 55$

$$I_C = h_{fe} \times I_b = 55 \times 3.488 = \mathbf{191.84mA}$$

Considering that I_C and I_b are known, U_{CE} can be determined using the diagram: $I_C = f(U_{CE})$ for I_b = Ctc



MJE 13005 $I_{CE} = f(U_{CE}) \mid I_b$ = Ctc

Using the diagram $I_C = f(U_{CE}) \mid I_b$ = Ctc:
for $I_b \approx 3.5mA$ and $I_C \approx 192$ mA: $\boldsymbol{U_{CE} \approx 40V}$
M ($U_{CE} = 40V$, $I_C = 192mA$) $\mid I_b = 3.5mA$

$$U_{in} \geq U_{CE} + U_{out} + U_{Z\,D1} - U_{bE} = 40 + 300 + 15 - 0.7 = 354.3V; \qquad U_{in} > 355V$$

$$R_1 = \frac{U_{in}-(U_{out}+U_{Z\,D1})}{I_{Z\,D1}+I_b} = \frac{355-(300+15)}{5+3.5} = \mathbf{4.7\ k\Omega}; \qquad R_1\text{: } 4.7\,k\Omega\,/\,2W$$

Power dissipation of the electronic components used

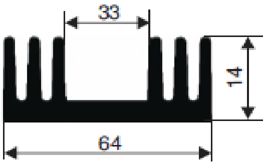
Power dissipation of LM317: 2W, heat sink:

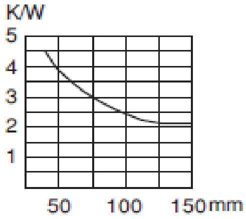
$$R_{\theta\,HA} \leq \frac{(T_J - T_A)}{P_d} - R_{\theta\,JC} - R_{\theta\,CH} = \frac{90-25}{2} - 5 - 1.2 = \mathbf{26.3^0C/W}$$

The LM317 can be mounted on a very small heat sink:

Thermal resistance: 16.7^{0}C/W
Length: 12.7 mm
Height: 29.9 mm
Width: 25.5 mm

Power dissipation of MJE 13005:

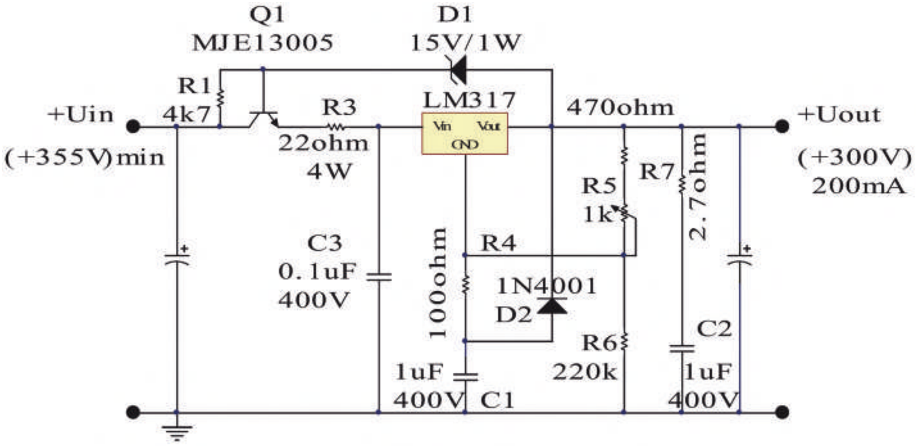




$$P_{d\,MJ13005} = U_{CE} \times I_C = 40V \times 191.84mA = 7.67W$$

$$\boldsymbol{R_{\theta\,HA}} \leq \frac{90-25}{7.67} - 1.67 - 1.2 = \boldsymbol{5.6^0C/W}$$

Heat sink length: (30mm ÷ 50mm)
Schematic:



Example:
High voltage regulator - Maida type: U_{out} = 380V, I_{out} = 200mA
*High voltage MOSFET **IRF 840**; 125W, 500V, 8A N – Channel MOSFET.*

The design procedure is similar to the design procedure of the BJT Maida type regulator.

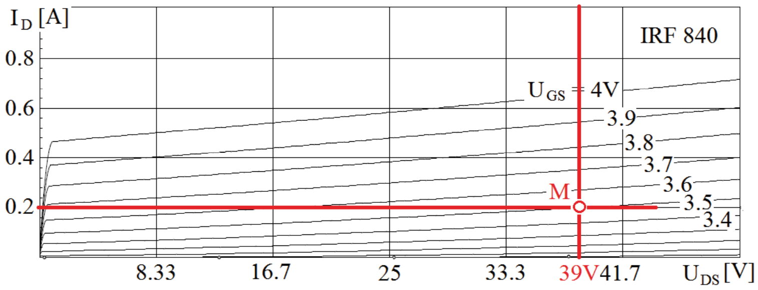
$$U_{Z\,D1} \geq U_{GS} + U_{in-out\,LM317} = 3.5V + 10V = 13.5V \quad (IRF840\!: U_{GS} \approx 3.5V)$$

*Zener diode: 15V / 1W (**$U_{Z\,D1-R}$ = 15V**)*

$$R_3 = \frac{U_{Z\,D1-R} - U_{Z\,D1}}{I_{out}} = \frac{15-13.5}{0.2} = 7.5\Omega; \quad \boldsymbol{R_3\,standard = 10\,\Omega}$$

R_6 standard = 220 kΩ / 4W

$$R_5 = \frac{R_6}{\frac{U_{out}}{1.25}-1} = \frac{220k\Omega}{\frac{380V}{1.25}-1} = 726\ \Omega; \quad \boldsymbol{R_5: 620\ \Omega + 250\ \Omega\ \ trim\ pot}$$



Using the diagram $I_D = f(U_{DS})\,|\,U_{GS}$ = Ctc for U_{GS} ≈ 3.5V and I_D ≈ 200 mA:
U_{DS} ≈ 39V

M (U_{DS} = 39V, I_D = 200mA) $|\,U_{GS}$ = 3.5V

$$\boldsymbol{U_{in}} \geq U_{DS} + U_{out} + U_{Z\,D1} - U_{GS}$$

$$= 39 + 380 + 15 + 3.5$$

$$= 430.5V$$

$$U_{in} > 431V$$

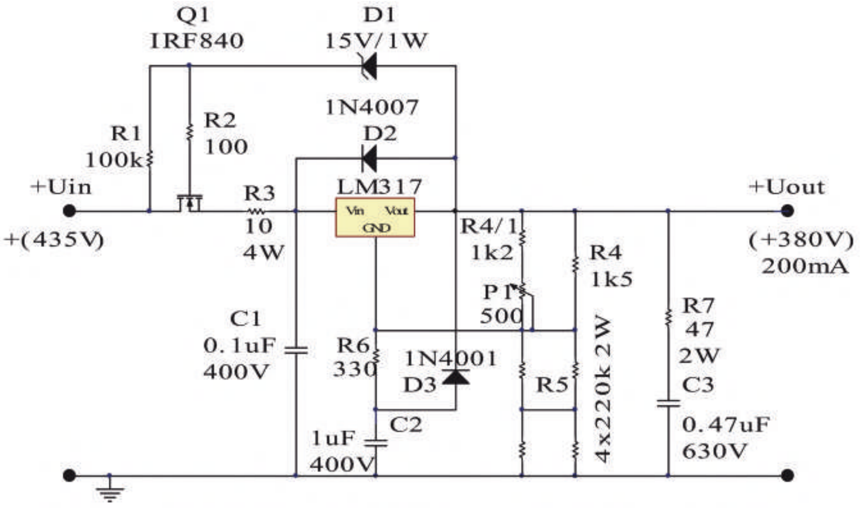
$P_{d\,IRF840} = U_{DS} \times I_D = 39V \times 200mA = 7.8W$ *(heat sink:* $R_{\theta\,HA} \approx 5^0C/W$*).*

Considering that the gate current of the MOSFET is very low (<100nA) and that the current of the Zener diode can be very low (<1mA), the resistor R_1 *can be very high resistance (100 kΩ).*

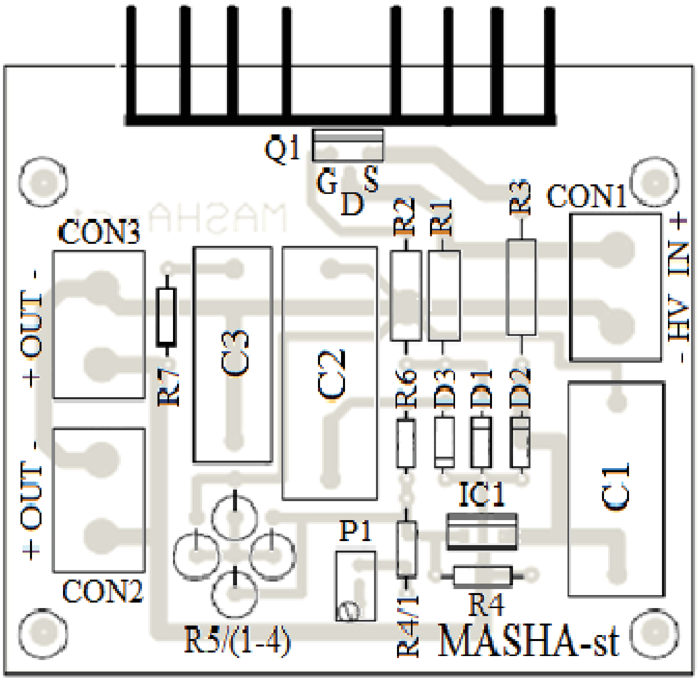
The power supply rejection ratio (ripple and noise) of the Maida high voltage regulator is very high (PSRR: over 110dB). Combined with the input CLC filter and the output capacitor, the PSRR is very high and the Maida regulator can be used as a power supply for low signal preamplifiers such as MM tube preamplifiers.

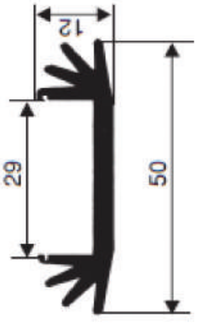
Even the better PSRR can be achieved by using the LT 3080 (adjustable 1.1A single resistor low dropout regulator) instead the LM317, as published by Mr. Tom Christiansen (Neurochrome).

Practical circuit of a *high voltage adjustable power supply -* ***MAIDA*** *type*
P1 - adjustment of the output voltage (U_{out} *).*

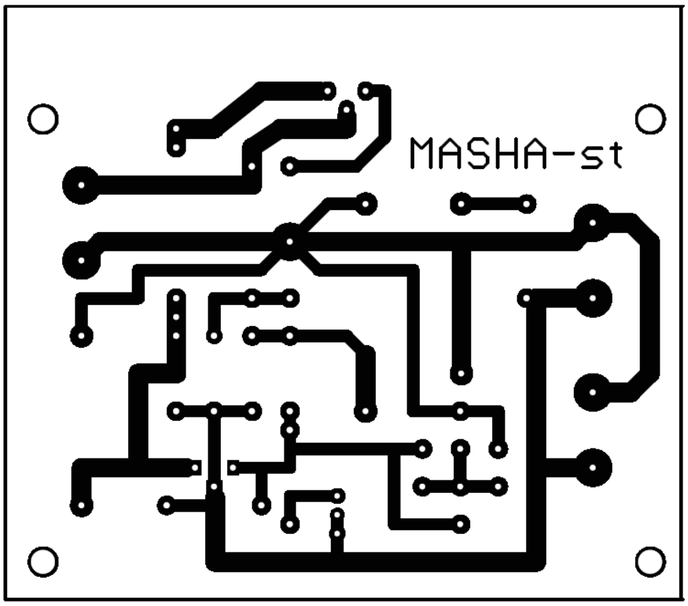


Note:
For $I_{out} \leq 200mA$
$U_{in} \approx U_{out} + 50V$
U_{in} = (400V … 450V)
U_{out} = (350 … 400V)



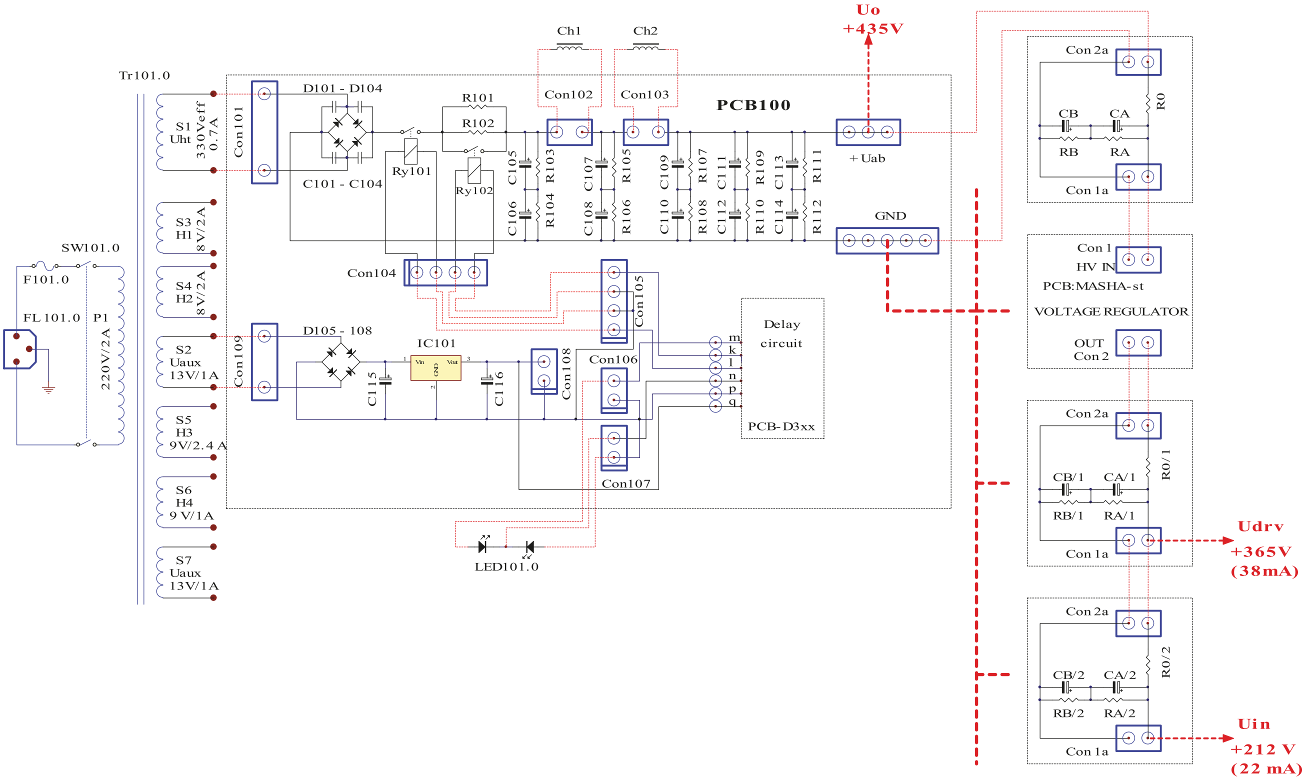


L = 30 mm



Cu side

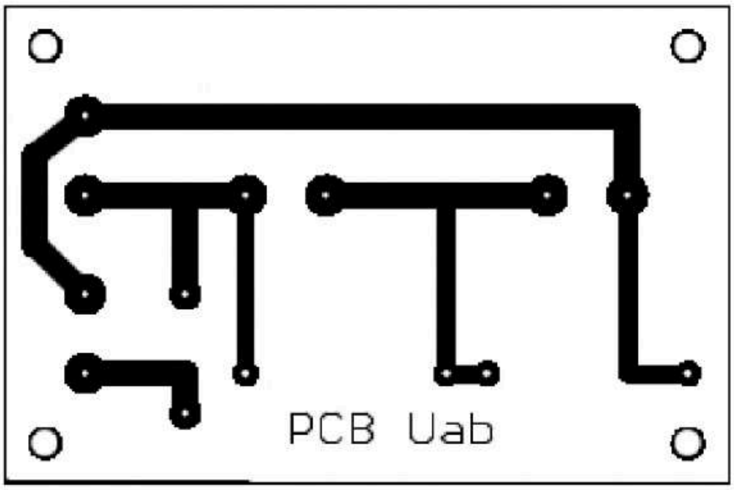
Example
High-End amplifier high-voltage power supply
Output stage: 435V / (162 ... 200) mA
Driver stage: 365V / 38mA
Input stage: 212V / 22mA

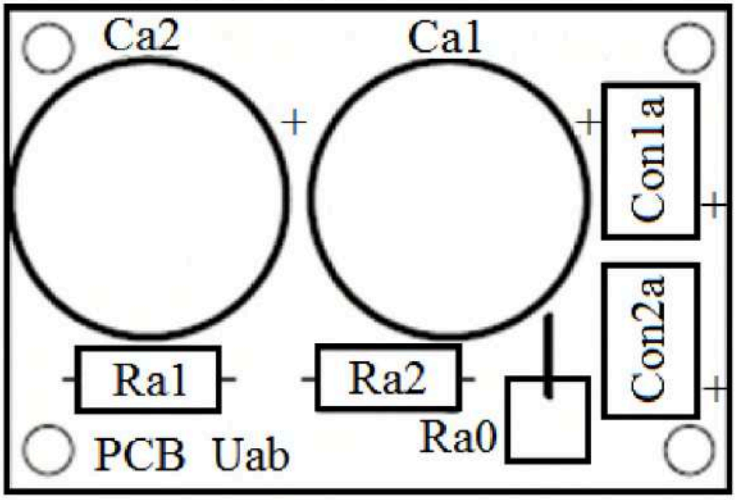


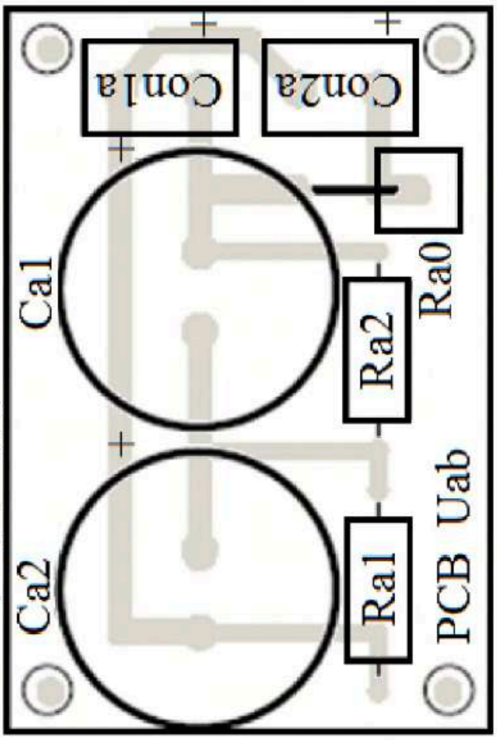
PCB 100

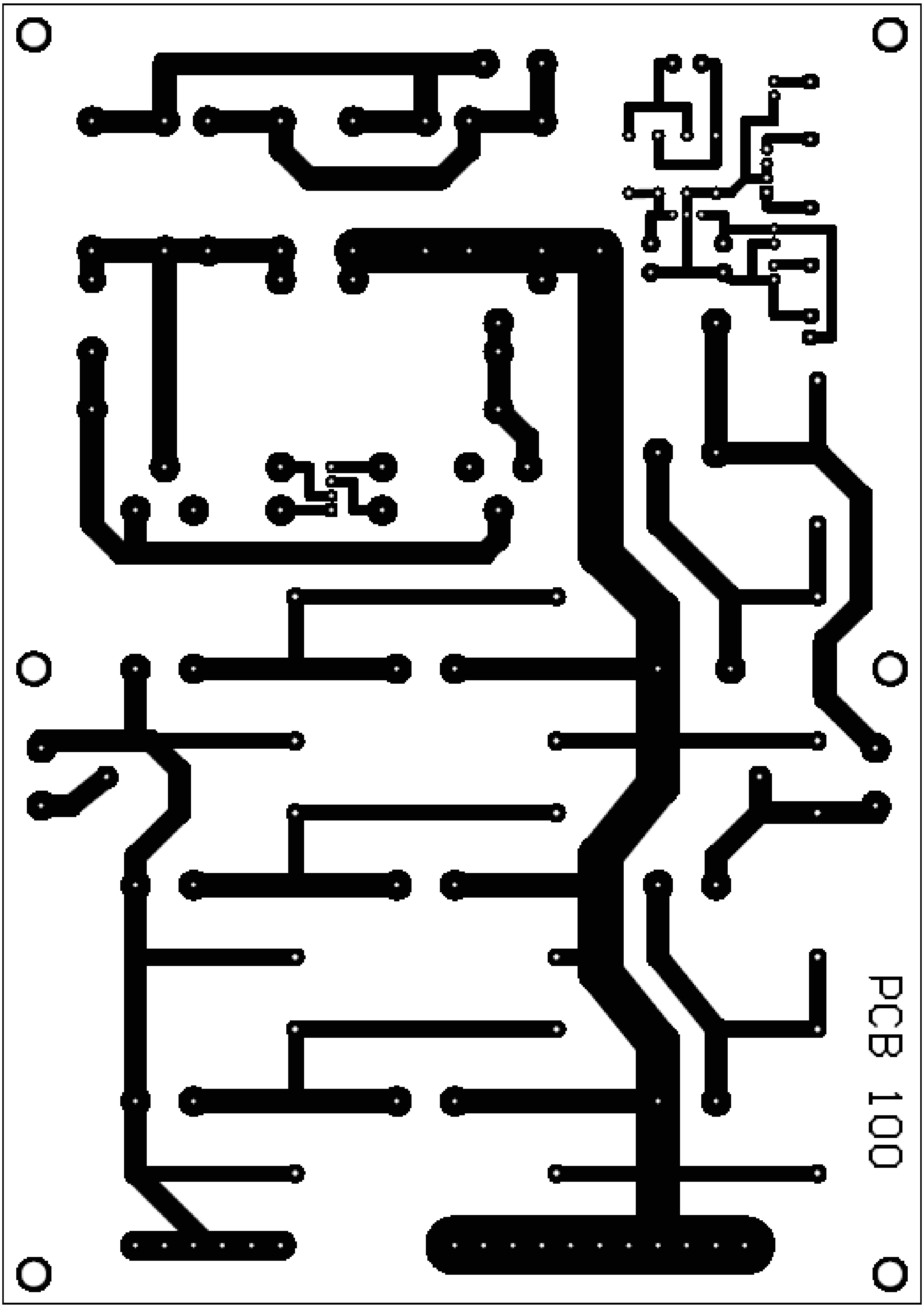
D101, D102, D103, D104 - BYT 56M
D105, D106, D107, D108 - 1N4002
C101, C102, C103, C104 - 100nF / 630V MKP
IC 101 - 7812 Stab.
LED 101.0 - RED - GREEN LED 5mm

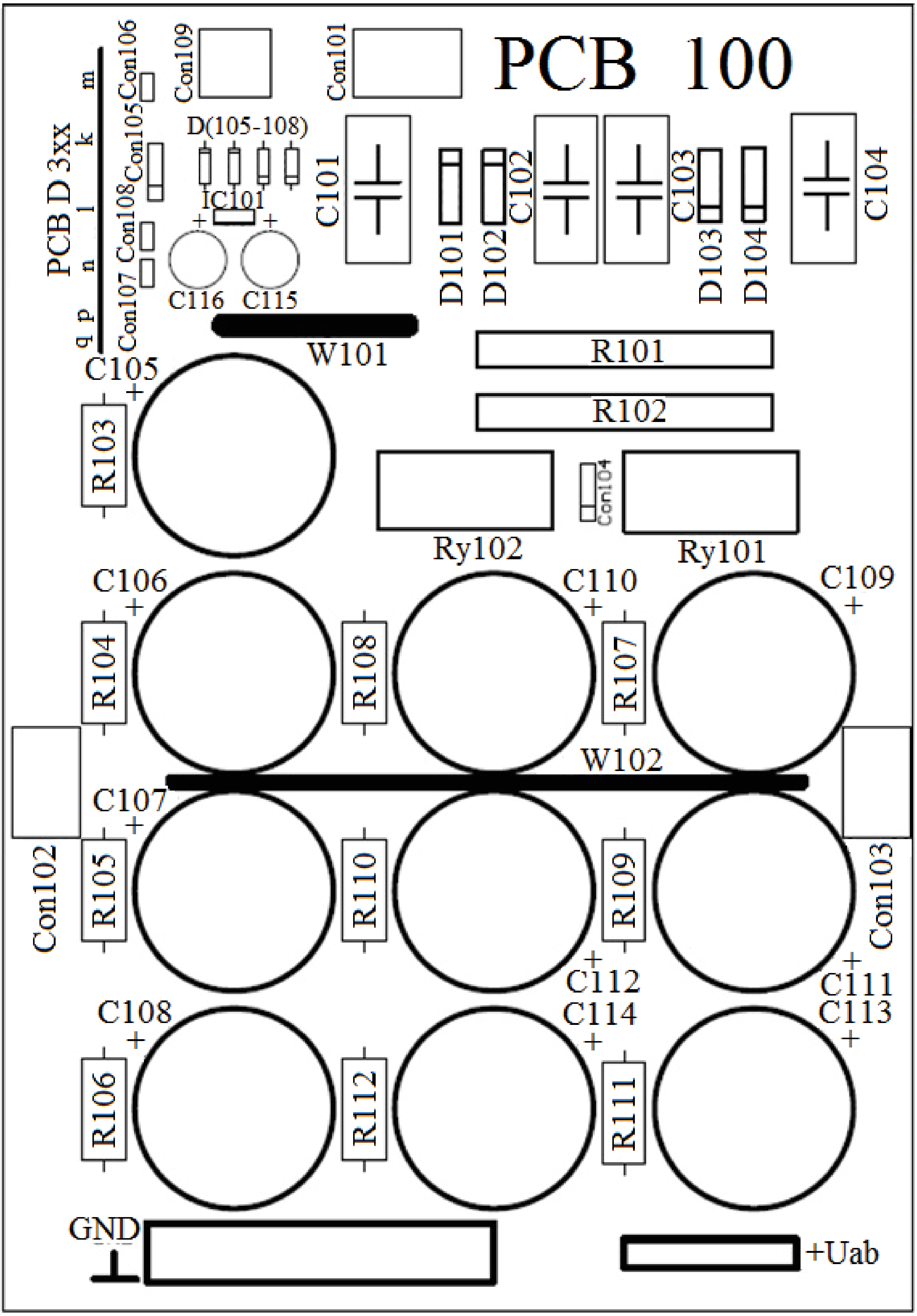
C105, C106, C107, C108, C109, C110, C111, C112, C113, C114, CA, CB. CA/1, CB/1 - 470 μF / 350V
CA/2, CB/2 - 330 μF / 350V
C115, C116 - 470 μF / 25V
R101, R102 - 1KΩ / 17W
R103, R104, R105, R106, R107, R108, R109, R110, R111, R112,
RA, RB, RA/1, RB/1, RA/2, RB/2 – 220kΩ/2W
R0 - 470 Ω / 11W
R0 / 1 - wire jumper
R0 / 2 - 6k8 / 17W
Ry 101, Ry 102 - 12V (250V, 15A)
Ch 1, Ch 2 - 5H / 250mA
PCB D 3xx - Delay circuit
PCB MASHA – St - 365V Maida voltage regulator
PCB Uab / 0, PCB Uab / 1, PCB Uab / 2 - PCB Uab (Filter capacitors PCBs)

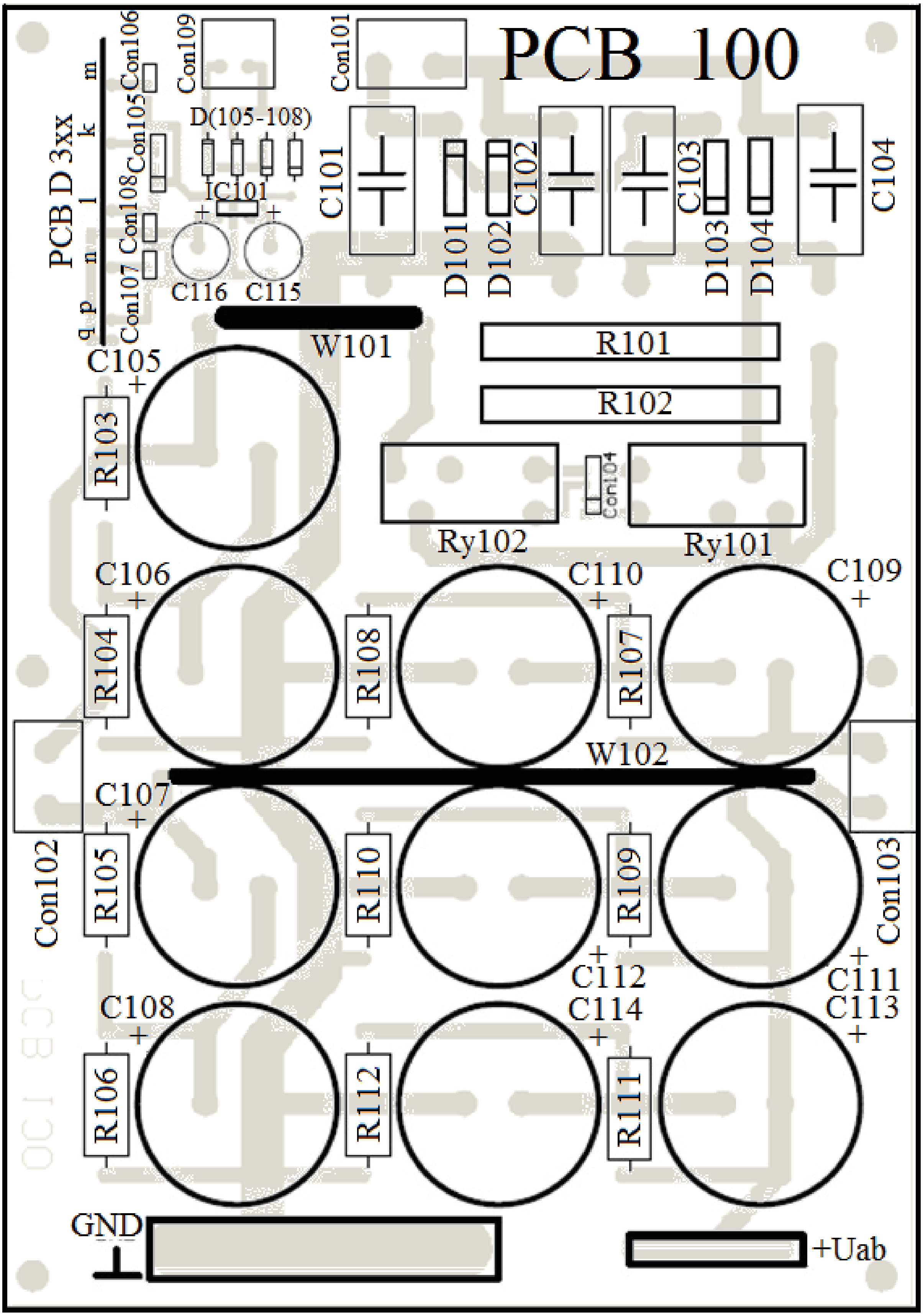


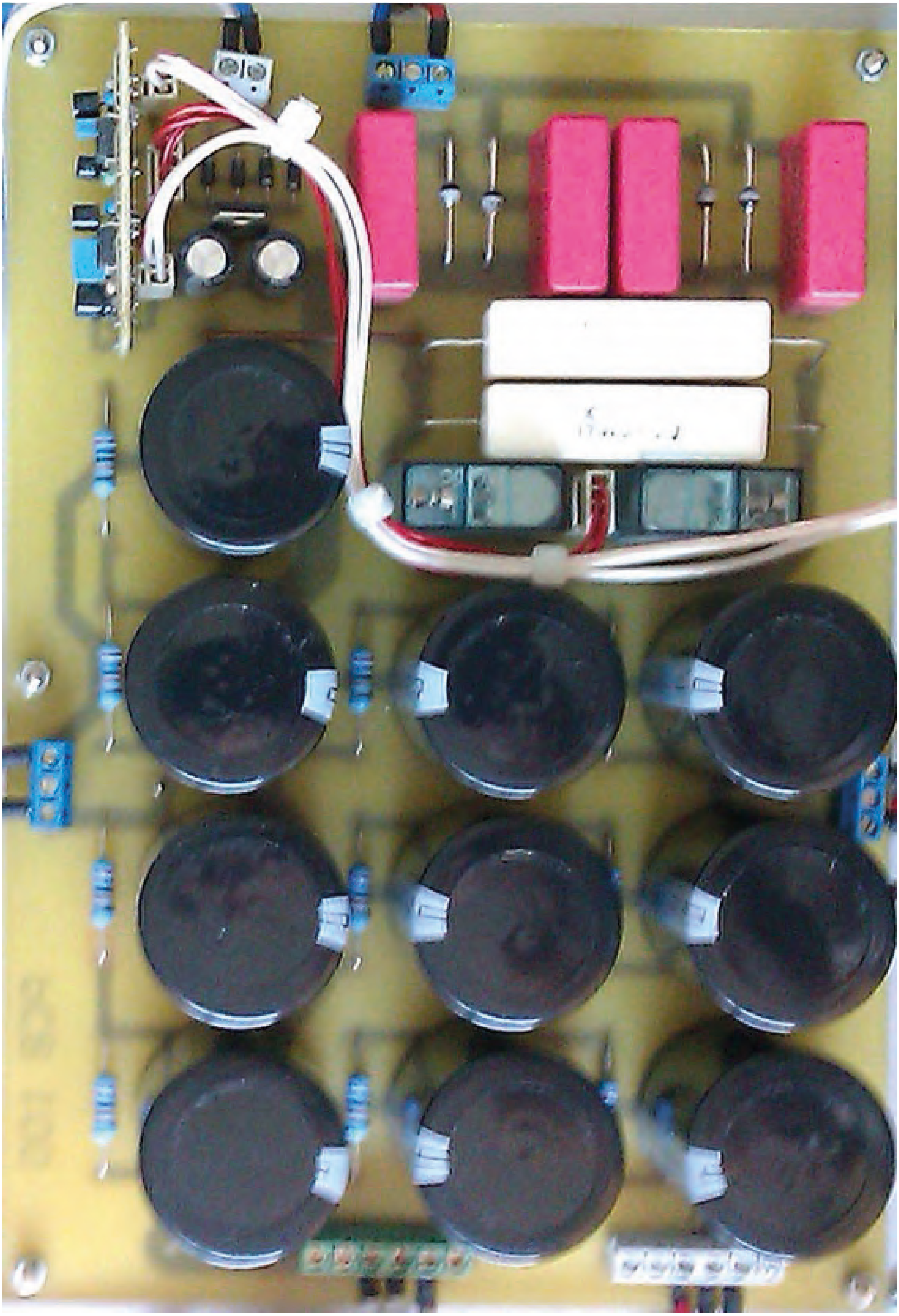




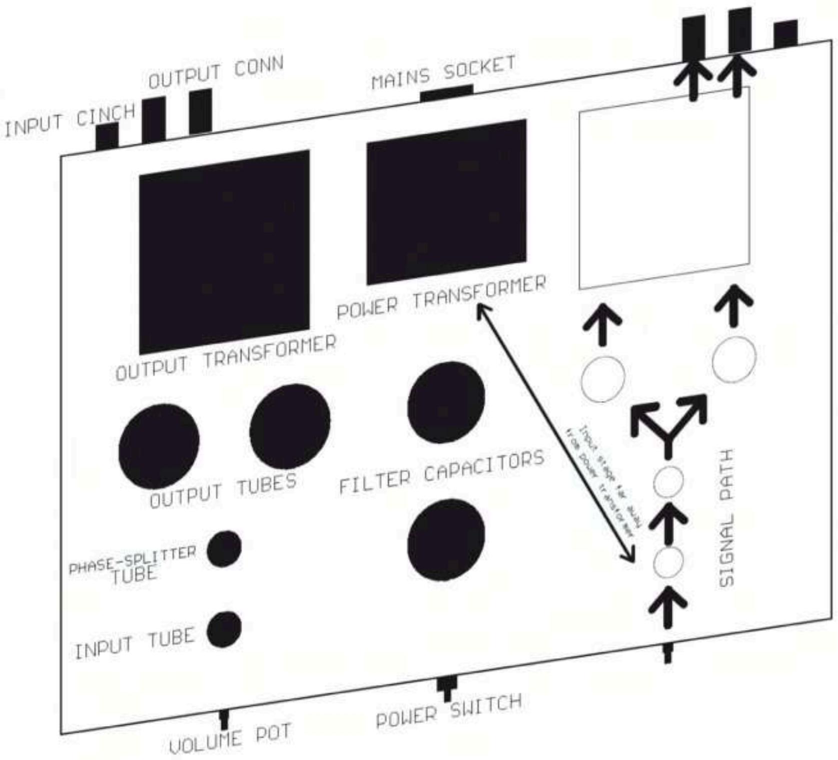

PCB 100


PCB 100
Con109
Con101
Con106
Con105
Con108
Con107
PCB D 3xx
m
k
l
n
p
q
D(105-108)
IC101
C101
C102
C103
C104
D101
D102
D103
D104
C116
C115
W101
R101
R102
C105
R103
Ry102
Ry101
Con104
C106
C110
C109
R104
R108
R107
W102
C107
R105
R110
R109
Con102
Con103
C112
C111
C108
C114
C113
R106
R112
R111
GND
+Uab


PCB 100
PCB D 3xx
m
k
l
n
q p
Con106
Con105
Con108
Con107
Con109
D(105-108)
IC101
C116
C115
Con101
C101
D101
D102
C102
C103
D103
D104
C104
W101
R101
R102
C105
R103
Ry102
Con104
Ry101
C106
R104
R108
C110
R107
C109
W102
C107
Con102
R105
R110
R109
Con103
C112
C111
C114
C113
C108
R106
R112
R111
GND
+Uab



Chapter 7 • In Practice



7.0 PRACTICAL POWER AMPLIFIER DESIGN

"At the heart of all ingenious things is their own simplicity".

Audiophile requirements:
High-End Class-A Vacuum Tube Power Amplifier for home music listening.

Procedure:
Decision: Single End or Push Pull amplifier.
Output power (taking into account the size of the room for listening to music, the efficiency of the loudspeaker, or just the wishes of the designer).

Block diagram (input, driver and output stage).
Choice of output tube.
Calculation of output stage (power, load – output transformer, voltage amplification, power supply - voltage and current).
Choice of input and driver tubes.
Calculation of input and driver stage (voltage amplification, power supply voltage and current).
Negative feedback.
Complete electric diagram of power amplifier / schematic.
Power supply.
Choice of electronic components.
Lay out of electronic and electromechanical components and chasses.
Mounting components on the chasses.
Wiring.
Testing and measurements.

Advice

The "mission" of the audio amplifier (and its designer) is to transform the input signal into an amplified output signal suitable for driving the loudspeaker and at a same time keep the output signal shape as similar as possible to the input signal shape (high fidelity).

 # The "Golden Rules" of designing

1. *The signal path from the input to the output of the amplifier has to be as* ***short*** *as possible, i.e., use only the necessary active and passive components to meet the project requirements. Each active and passive component also has some negative effects to the signal processing (technological limitation in the production of ideal passive or active components - spreading and non linearity of electrical characteristics, changes in characteristics caused by temperature fluctuation, noise…).*

2. *"Worst case" designing methodology*
   In the calculation, use the worst parameters and values of components within the tolerance field published in their data sheets – minimizing the risk that the performance of the actual construction will be worse than the performance obtained by the calculation.

3. *Standard values of electronic components*
   After calculating the value of the electronic components for further calculation, use the nearest standard values of the electronic components in accordance with the above "worst case" rule.

4. *Components from the current production*
   To ensure that the design does not remain just a design on paper, use components from current and stable production.

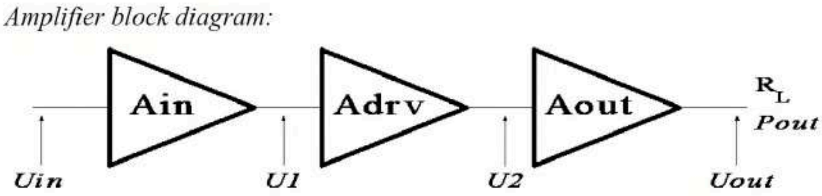
5. *The operating condition of the components must be within the min. to max. datasheet values.*
   Max. designed values of operating conditions must be at least (10 – 20) % lower than max. values published in data sheets.

Example:
Low power amplifier suitable for listening to music in the home ambience or small apartments with an average size of 16 ÷ 20 m^2 *and using loudspeaker boxes with an efficiency of 90 ÷ 93 dB.*

Design requirements:
Single-Ended amplifier, output power (5 … 8)W, R_L = 4Ω,
U_{in} ≤ 250mV$_{RMS}$ (or, U_{in} ≤ 780mV$_{RMS}$ with applied NFB).

Amplifier block diagram:

Amplifier block diagram:



Total voltage amplification of the amplifier:

$$A_{Tot} = \frac{U_{out}}{U_{in}} = \frac{\sqrt{P_{out} \times R_L}}{U_{in}}$$

$$A_{Tot} = A_{in} \times A_{drv} \times A_{out}$$

If the input voltage is U_{in} = 250mV and P_{out} ≈ 7W at a load of 4Ω, the total voltage amplification must be:

$$A_{Tot} \geq \frac{U_{out}}{U_{in}} = \frac{\sqrt{P_{out} \times R_L}}{U_{in}} = \frac{\sqrt{7W \times 4\Omega}}{0.25V} \geq \mathbf{21.6\ x}\ or\ \ \mathbf{26.69dB}$$

Amplifier design starts with the design and calculation of the output stage.
The first choice for the output tube is the unsurpassed WE 300B directly heated triode (audiophile reference standard for good sound).

Output stage is already explained elsewhere in this book (optimum quiescent point, load line…):
WE 300B, Quiescent point:
M (U_{ak} = 350V, I_a = 80mA), U_{gk} = - 70V, Ra = 3500 Ω.
Heater power supply (DC): 5V$_{DC}$, 1.3A.

Calculation of output stage voltage amplification:

$$\mathbf{A_{out}} = \mu_{300B} \times n_{OT} = 3.2 \times \sqrt{\frac{R_{Load}}{R_a}} = 3.2 \times \sqrt{\frac{4\Omega}{3500\Omega}} = \mathbf{0.108}\ x,\ \ or\ \ -\mathbf{19.317}\ dB$$

n_{OT} – *turns ratio of the output transformer* $n = \frac{N_S}{N_P} = \sqrt{\frac{R_{Load}}{R_a}}$

Expected output power: $P_{out} = \frac{(U_{in\ o\ RMS} \times A_{out})^2}{R_{Load}} = \frac{\left(\frac{U_{in\ Peak}}{\sqrt{2}} \times A_{out}\right)^2}{R_{Load}} = \frac{\left(\frac{70V}{\sqrt{2}} \times 0.108\right)^2}{4\Omega} = 7.18W$

Necessary driving voltage: U_2 = 70V$_{Peak}$ = 49.6V$_{RMS}$ ≈ 49V$_{RMS}$

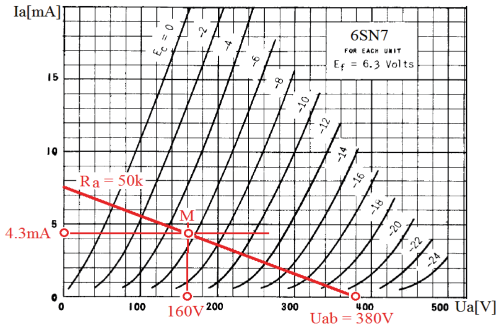
The next step is: calculating the total voltage amplification of the input and driver stage:

$$A_{in} \times A_{drv} \geq \frac{A_{Tot}}{A_{out}} = \frac{21.6}{0.108} \geq 200\ x\ \ or,$$

$$A_{in}[dB] + A_{drv}[dB] = A_{Tot}[dB] - A_{out}[dB] = 26.69 - (-19.317) \geq 46dB$$

As the total voltage amplification of the input and driver stage is known (200 x), the design of the amplifier can be continued by selecting the driver and input stage tubes. The voltage amplification of the input and driver stage can be obtained by using small signal triodes with lower μ (estimation: if the amplification is evenly distributed on both stages, it is sufficient that the gain per stage is $\sqrt{200} \approx 14\ x$ *. The well-known 6SN7double triode (μ = 20) is one of the first choices for driver and input stage.*

Driver stage: classic grounded cathode amplifier (1 / 2 6SN7):



The condition that the output signal of the driver tube is 140 $V_{P\text{-}P}$ as well as the additional condition that the power supply of the driver and input stage does not exceed 420 V_{DC} (common power supply circuit is used for output, driver and input stage with all filters or voltage regulator circuits - driver stage power supply of 380V is chosen) is satisfied with the correct choice of operating conditions of the tube 6sn7, i.e. the choice of the quiescent point and the load line.

Quiescent point:
M (160V, 4.3 mA) $|U_g = -5.2$ V and power supply of 380 V_{DC} .
$R_a = 50\ k\Omega$

Voltage amplification of the driver stage:

$$A_{drv} = \mu \times \frac{R_a}{R_i+R_a} = 20 \times \frac{50k\Omega}{50k\Omega+10k\Omega} = \mathbf{16.6}\ x\ \ or,\ \ \mathbf{24.4}\ dB$$

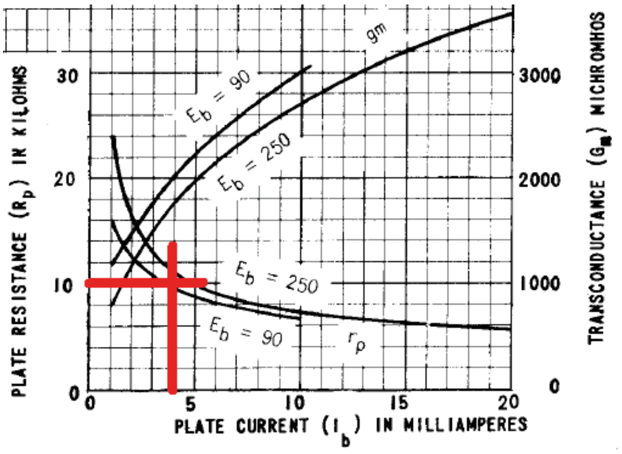
Internal resistance of 6SN7, $R_i \approx 10k\Omega$ at $U_a = 160$ V, $I_a = 4.3mA$ and $U_g = -5.2$ V

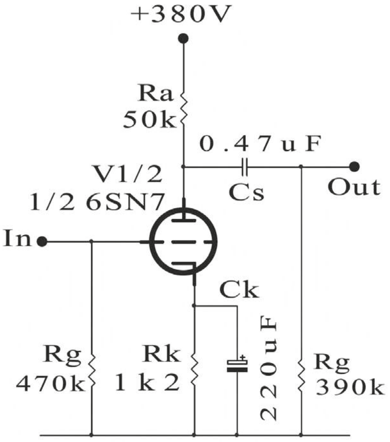
Cathode resistor: $R_{k\ drv} = \frac{U_g}{I_a} = \frac{5.2V}{4.3mA} = 1.2k$

Bypass capacitor of the cathode resistor: $C_k = \frac{1}{2\times\pi\times R_{keq}\times f_{L\,(-3dB)}}$, $R_{k\ eq} = R_k' \| R_k = \frac{\frac{R_i+R_a}{\mu+1}\times R_k}{\frac{R_i+R_a}{\mu+1}+R_k}$, $R_k' = \frac{R_i+R_a}{\mu+1}$

$$R_k' = \frac{R_i+R_a}{\mu+1} = \frac{10+5}{21} = 2.857k\Omega,\ \ R_{k\ eq} = R_k' \| R_k = 2.857\,|\,|\,1.2 = 845\Omega\,,$$

$$For\ f_{L\,(-3dB)} = 2Hz:\ \ C_k \geq \frac{1}{2\times\pi\times R_{keq}\times f_{L\,(-3dB)}} = \frac{1}{2\times3.14\times845\times2} \geq 94.22\mu F,\ \ C_k = 220\mu F\,/\,25V$$





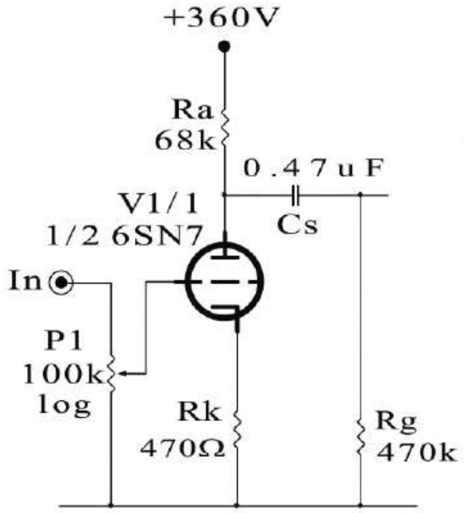
Calculation of input stage voltage amplification:

$$A_{in} \geq \frac{A_{Tot}}{A_{out} \times A_{drv}} = \frac{21.6}{0.108 \times 16.6} \geq \mathbf{12} \times \ \ or,$$

$$A_{in}[dB] = A_{Tot}[dB] - A_{drv}[dB] - A_{out}[dB] = 26.69 - 24.4 - (-19.317) \geq \mathbf{21.6\ dB}$$

Input stage: classic grounded cathode amplifier (1 / 2 6SN7):

Such a circuit is already designed elsewhere in this book.



Voltage amplification of the input stage:

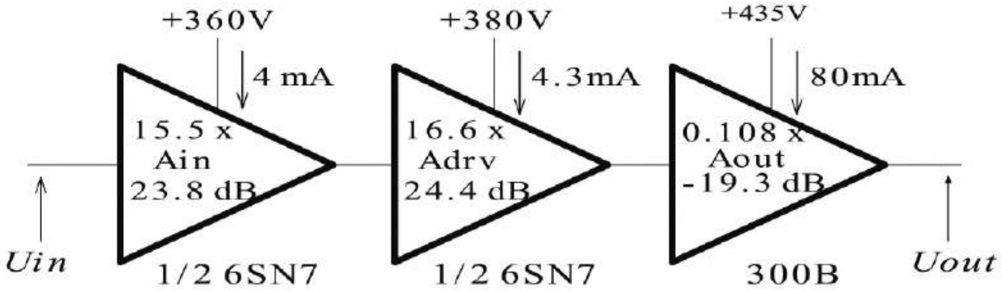
Quiescent point M (90V, 4mA)│U_g = – 1.9 V

Cathode resistor: $R_{k\,in} = \frac{U_g}{I_a} = \frac{1.9V}{4mA} = 475\,\Omega$, $\boldsymbol{R_{k\,in} = 470\,\Omega}$

$$A_0 = -\mu \times \frac{R_a}{R_a + R_i} = -20 \times \frac{68k}{68k + 10k} = -17.4$$

$$A_{in} = \frac{A_0}{1 + \frac{R_k}{R_a} \times A_0} = \frac{17.4}{1 + \frac{0.47}{68} \times 17.4} = \mathbf{15.5}\ or, \qquad \mathbf{23.8}\,dB$$

A complete block diagram:



Total voltage amplification of the designed amplifier:

$$A_{Tot} = \frac{U_{out}}{U_{in}} = A_{in} \times A_{drv} \times A_{out}$$

$$\boldsymbol{A_{Tot}} = 15.5 \times 16.6 \times 0.108 = \mathbf{27.78} \times$$

$$A_{Tot}[dB] = A_{in} + A_{drv} + A_{out}$$

$$\boldsymbol{A_{Tot}}[dB] = 23.8 + 24.4 - 19.3 = \mathbf{28.9}\,dB$$

Input voltage required for full power of the amplifier (7 W) – amplifier input sensitivity:

$$\boldsymbol{U_{in}} = \frac{U_{out}}{A_{Tot}} = \frac{\sqrt{P_{out} \times R_L}}{A_{Tot}} = \frac{\sqrt{7W \times 4\Omega}}{27.78} = \mathbf{190}\,mV$$

Coupling capacitors, input stage – driver stage, driver stage – output stage: C_S = 0.47 μF / 400 *V.*

Before synthesizing the final electric schematic, it is necessary to determine the additional characteristics of the electronic components (dissipation of resistors and their rated power, rated voltage of capacitors …).

Resistors:
Cathode resistor of the input stage tube R_k (denoted as R1on the final electric schematic):
***R1**: Dissipation $P_{d\,R1}$ = R1 × I_a^2 = 0.47 kΩ × 4.0^2 mA ≈ 7.5 mW; R1 = **470 Ω / 1 W** Metal film*

Anode resistor of the input stage tube R_a (denoted as R2 on the final electric schematic):
***R2**: Dissipation $P_{d\,R2}$ = R2 × I_a^2 = 68 kΩ × (4.0 mA)^2 ≈ 1.096 W*

*The **rated power** of the resistor must be **4 to 5 times** higher than the calculated dissipation of the resistor.*
$P_{R\,rated}$ = (4 ÷ 5) × $P_{d\,R\,calculated}$

It is hard to find a metal film resistor with a power of more than 2 W. Therefore, a combination of several 2 W metal film resistors can be used (in parallel):
R2**:* ***220 *kΩ / 2 W || **220*** *kΩ / 2 W || **180*** *kΩ / 2 W = **68.27*** *k Ω* ***/ 6*** *W*
Three 220 kΩ / 2 W metal film resistors connected in parallel.

Grid resistor of the driver stage tube
R3**: **470 *kΩ **/ 1 W** metal film*

Cathode resistor of the driver stage tube R_k (denoted as R4 on the final electric schematic):
***R4**: Dissipation $P_{d\ R4} = R4 \times I_a^2 = 1k2 \times (4.3\ mA)^2 \approx 22\ mW$; R4 = **1k2 / 1 W** Metal film*

Anode resistor of the driver stage tube R_a (denoted as R5 on the final electric schematic):
***R5**: Dissipation $P_{d\ R5} = R5 \times I_a^2 = 50\ k\Omega \times (4.3\ mA)^2 \approx 0.925\ W$*
R5**: Three 150k Ω / 2W metal film resistors connected in parallel – **50 *kΩ **/ 6 W***

Grid resistor of the output stage tube
R6**: **390 *kΩ **/ 1 W** Metal film*

Cathode resistor of the 300B tube (denoted as R7 on the final electric schematic):
***R7**: $P_d = R_7 \times I_a^2 = 880\ \Omega \times 0.08\ A^2 = 5.632\ W$,*
*Standard: **R7 = 900*** *Ω **/ 30 W**, (**3 × 2k7 / 10*** *W – in parallel connection , wire wound resistors)*

R8, R6**: **33 *Ω **/ 2**W Metal film*

P1**: **100 *kΩ **/ log***

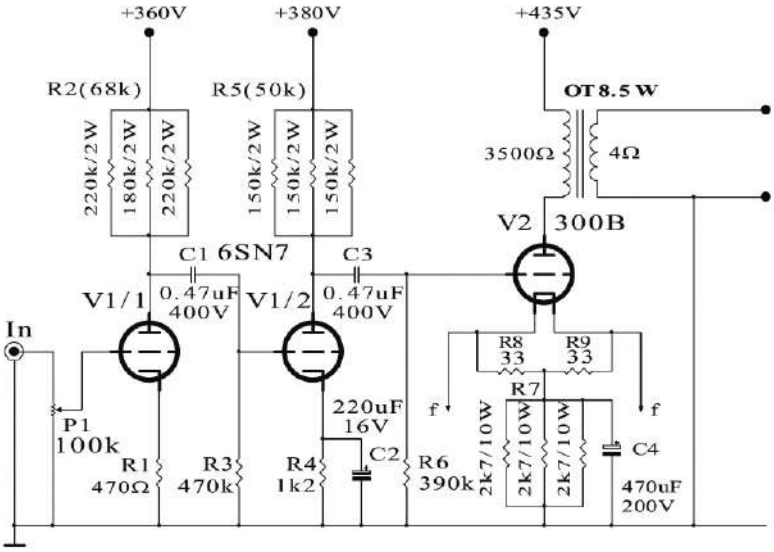
Capacitors:
Bypass capacitor of the cathode resistor of the driver stage tube (denoted as C2 on the final electric schematic):
C2**: Voltage across the R3 is around 5.2V, C2 = (100 ÷ 470) μF / 16V ; **C2 = 220 *μF **/ 16*** *V*

Coupling capacitors (denoted as C1 and C3 on the final electric schematic):
C1, C3**: High quality audio coupling capacitor. Foil MKP or other **0.47 *μF **/ 400*** *V*

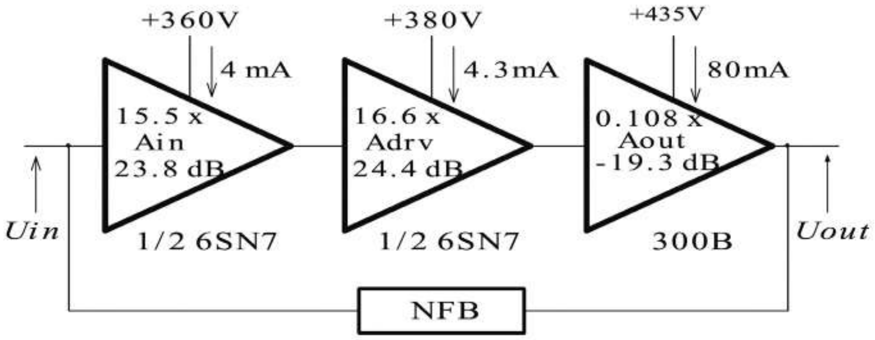
Bypass capacitor of the cathode resistor of the output stage tube (denoted as C4 on the final electric schematic):
C4**: **470 *μF **/ 200*** *V (rated voltage: more than twice the voltage across the R6)*

OT**: Output transformer **3500 *Ω **/ 4*** *Ω **/ 8.5*** *W, $\mathbf{R_{Primary - DC} = 188.5}\ \Omega$*
** DC resistance of the Primary plays a role in the calculation of the high voltage power supply of the output stage:*
$U_{ab\ out} = U_{Cathode\ 300B} + U_{a\text{-}c\ 300B} + R_{Primary - DC} \times I_{a\ 300B}$.

Complete electric schematic of the amplifier:



Amplifier characteristics related to some main characteristics such as total harmonic distortion and amplitude characteristic can be improved by adding negative feedback to the amplifier (for a Single-End amplifier, the use of NFB up to –10dB is allowed):



The voltage amplification of the amplifier with applied NFB is lower (by the amount of NFB) than the amplification without applied NFB:
With an applied NFB of ***– 10 dB*** *(**3.16 x**), the amplification is:* $\boldsymbol{A_{Tot\,NFB}} = A_{Tot} - A_{NFB} = 28.9 - 10 =$ ***18.9 dB or 8.8***×
The input voltage required for the full power of the amplifier (input sensitivity of the amplifier) is higher by the amount of applied NFB:

$$\boldsymbol{U_{in\,T}} = U_{in} \times 3.16 = 0.19 \times 3.16 \approx \mathbf{600\ mV}$$

(Total harmonic distortion is 3.16 times less than the THD of the amplifier without applied NFB)
The NFB circuit can be applied between the output of the amplifier and the cathode circuit of the input tube, for example.

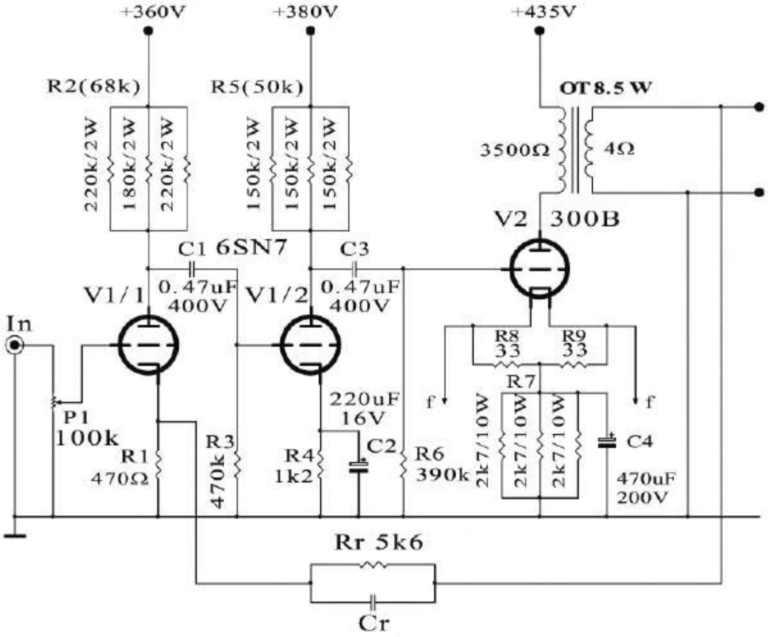
NFB circuit resistance R_r *calculation:* $A_{Tot} = 27.78\times$; $A_{Tot\,NFB} \approx 8.8\times$; $\mathbf{R1} = \mathbf{470\ \Omega}$

$$A_{Tot\,NFB} = 8.8 = \frac{A_{Tot}}{1+\beta_r \times A_{Tot}} = \frac{A_{Tot}}{1+\frac{R1}{R1+R_r} \times A_{Tot}} \rightarrow$$

$$R_r = R1 \times \left(\frac{A_{Tot} \times A_{Tot\,NFB}}{A_{Tot} - A_{Tot\,NFB}} - 1\right)$$

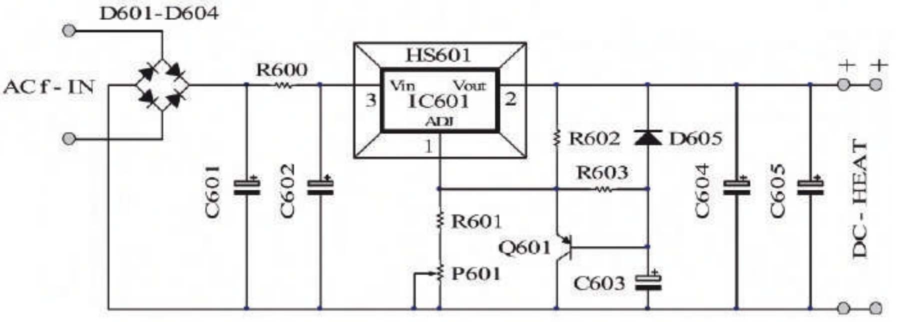
$$\boldsymbol{R_r} = R1 \times \left(\frac{A_{Tot} \times A_{Tot\,NFB}}{A_{Tot} - A_{Tot\,NFB}} - 1\right) = 470 \times \left(\frac{27.78 \times 8.8}{27.78 - 8.8} - 1\right) = 5583\,\Omega\,;\ \boldsymbol{R_r = 5k6}$$

Complete electric schematic of the amplifier with applied NFB:



Power supply
Heater power supply; DC type
6SN7: U_f = 6.3 V, 0.6 A
300B: U_f = 5 V, 1.2 A
The voltage regulator circuit using the LM338 can be applied.

For two channels it is necessary to use:
Two 5 V/1.2 A voltage regulator to supply the output tube heaters and one 6.3 V/1.2 A voltage regulator to supply the input and driver tube heaters.

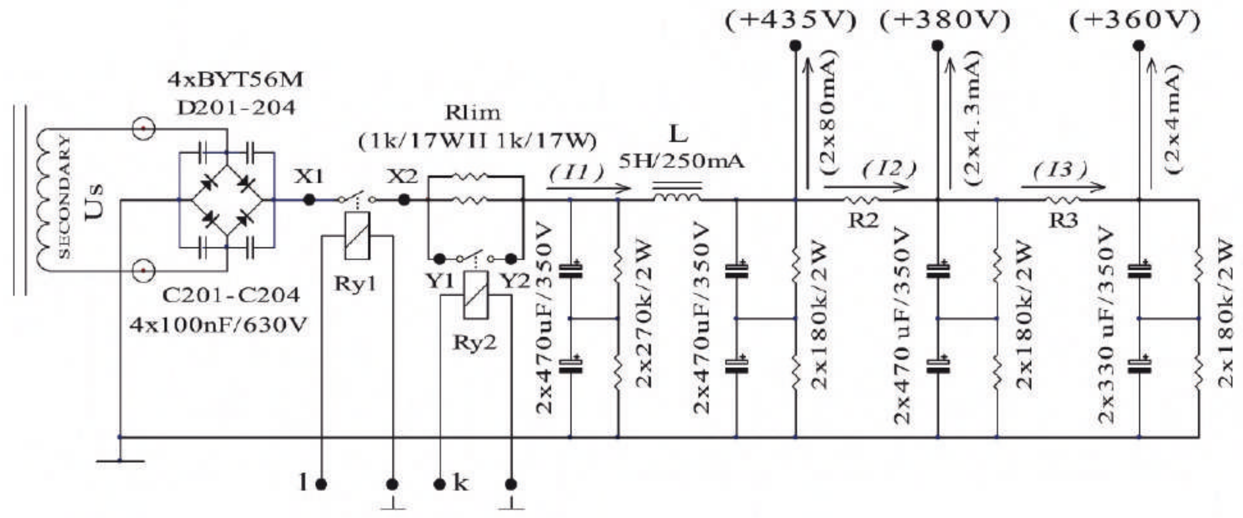


C601, C602: 10000 μF / 25 V
C604, C605: 4700 μF / 16 V
C603 – 100 μF / 16 V
R603 – 51 kΩ / 0.25 W
D605 – 1N4002
Q601 – BC557B
R602 – 120 Ω / 0.25 W
R601 – 270 Ω
IC 601 – LM338
D601 ÷ D604 – 1N5822

5V / 1.2A: R600 – 1 Ω / 10 W, P601 – 200 Ω trimpot., Secondary: 9 V_{RMS} / 1.8 A
6.3V / 1.2A: R600 – 0.5 Ω / 10 W, P601 – 470 Ω trim pot., Secondary: 10 V_{RMS} / 1.8 A
The LM338 must be mounted on a 5 K/W heat sink
PCB MM / 600 can be used, design details can be found in the text above.

High voltage power supply
*A **well filtered** high voltage power supply is one of the most important requirements that must be met when designing a single ended amplifier.*

Output stage: 435 V / 80 mA (one channel)
Driver stage: 380 V / 4.3 mA (one channel)
Input stage: 360 V / 4 mA (one channel)
Bridge rectifier with CLC filter



Calculating R2 and R3

R3: Voltage across R3 is 380 V – 360 V = 20 V
Current I3 is 2 × 4 mA + current flows through the resistors connected across the filter capacitors:
360V / (2 × 180 k) = 1 mA
I3 = 2 × 4 mA +1 mA = 9 mA and R3 is 20 V / 9 mA =2k2. R3 power dissipation is R3 × $I3^2$ *≈ 0.18 W*
R3 = 2k2 / 2 W

R2: Voltage across R2 is 435 V – 380 V = 55 V
Current I2 is 2 × 4.3 mA + I3 + current flows through the resistors connected across the filter capacitors:
380 V / (2 × 180 k) = 1.06 mA
I2 = 8.6 mA + 9 mA + 1.06 mA = 18.66 mA and R2 is: 55 V / 18.66 mA = 2k95. R2 standard = 2k7
R2 power dissipation is R2 × $I3^2$ *≈ 0.94 W*
R2 = 2k7 / 5 W

Total high voltage power supply current (I1):
I1 = (2 × 80 mA) +I2 + 2 × current flows through the resistors connected across the filter capacitors:
435 V / (2 × 180k) ≈ 1.21mA
I1 = 160 mA + 18.66 mA + 2.42 ≈ 181.1 mA

Power supply load is R_{Load} *= 435 V / 181.1 mA ≈ 2400 Ω*
Filter capacitor and filter choke can be calculated based on the required ripple voltage as explained in the power supply chapter of this book.

For filter capacitor (235 μF) and choke 5 *H / 250 mA /* $R_{choke-DC}$ *= 55 Ω (commonly), the expected ripple factor is:*

$$RF = \frac{\sqrt{2}}{8 \times (2 \times \pi \times f)^3 \times L \times C_1 \times C_2 \times R_{Load}} = \frac{\sqrt{2}}{8 \times (2 \times \pi \times 50)^3 \times 5 \times 235 \times 10^{-6} \times 235 \times 10^{-6} \times 2700} \approx \mathbf{8.6 \times 10^{-6}}$$

Ripple voltage (RMS) is:

$$U_{r\,(RMS)} = RF \times U_{DC} = RF \times U_0 = 8.6 \times 10^{-6} \times 435V \approx 3.74\,mV$$

Ripple voltage (U_1 *= 380 V and I 2 = 18.66 mA):*
AC voltage source of 3.74 mV_{RMS} *(ripple voltage at* U_0 *) and a voltage divider consisting of R2 and* C_2 *= 235μF .*
Impedance of C_2 *capacitor is:* $X_{C2} = 1 / (2\pi f C_2) = 1 / (2x3.14x100x235x10^{-6}) \approx 6.77\ \Omega$ *(f = 100Hz)*

$$U_{r\,(RMS)} = 3.74mV \times \frac{X_{C2}}{X_{C2}+R2} = 3.74mV \times \frac{6.77}{6.77+2700} \approx 0.009\ mV$$

Ripple voltage (U_2 *= 360 V and I 3 = 9 mA):*
AC voltage source of 0.009 mV_{RMS} *(ripple voltage at* U_1 *) and a voltage divider consisting of R3 and* C_3 *= 165* **μF .**

Impedance of C_3 *capacitor is:* $X_{C2} = 1 / (2\pi f C_2) = 1 / (2x3.14x100x165x10^{-6}) \approx 9.65\ \Omega$ *(f = 100 Hz)*

$$U_{r\,(RMS)} = 0.009mV \times \frac{X_{c3}}{X_{C3} + R3} = 0.009mV \times \frac{9.65}{9.65 + 2200} \approx 0.039\ \mu V$$

Voltage drop across the choke:
$U_{Choke} = R_{choke-DC} \times$ *total power supply current* $= 55\ \Omega \times 0.181\ A = 9.96\ V$

The rectified DC voltage (DC voltage at the output of the rectifier) is: $U_0 + 9.96\ V = 435\ V + 9.96\ V \approx$ ***445 V***

Secondary voltage calculation:

If long-life (LL) grade aluminum electrolytic capacitors are used:

$$\boldsymbol{U_{S\,[RMS]}} = \frac{U_{out} + I_{Load} \times (R_{S[DC]} + n^2 \times R_{P[DC]}) + 2 \times U_D}{\sqrt{2}}$$

*Power transformer, resistance of the Primary**:*** $R_P \approx 5\ \Omega$,
*Power transformer, resistance of the Secondary**:*** $R_S = 15\ \Omega$,
Ratio of Primary to high voltage Secondary: $n \approx 1.5$,
Forward voltage of the rectifier diode: $U_D \approx 0.8\ V$,

$$\boldsymbol{U_{S\,[RMS]}} = \frac{445 + 0.181 \times (15 + 1.5^2 \times 5) + 2 \times 0.8}{\sqrt{2}} = 320.1V \approx \mathbf{320\ V}$$

Secondary current must be: $\boldsymbol{I_{S\,[RMS]} \approx 1.5 \times I_{DC\,[Load]}}$

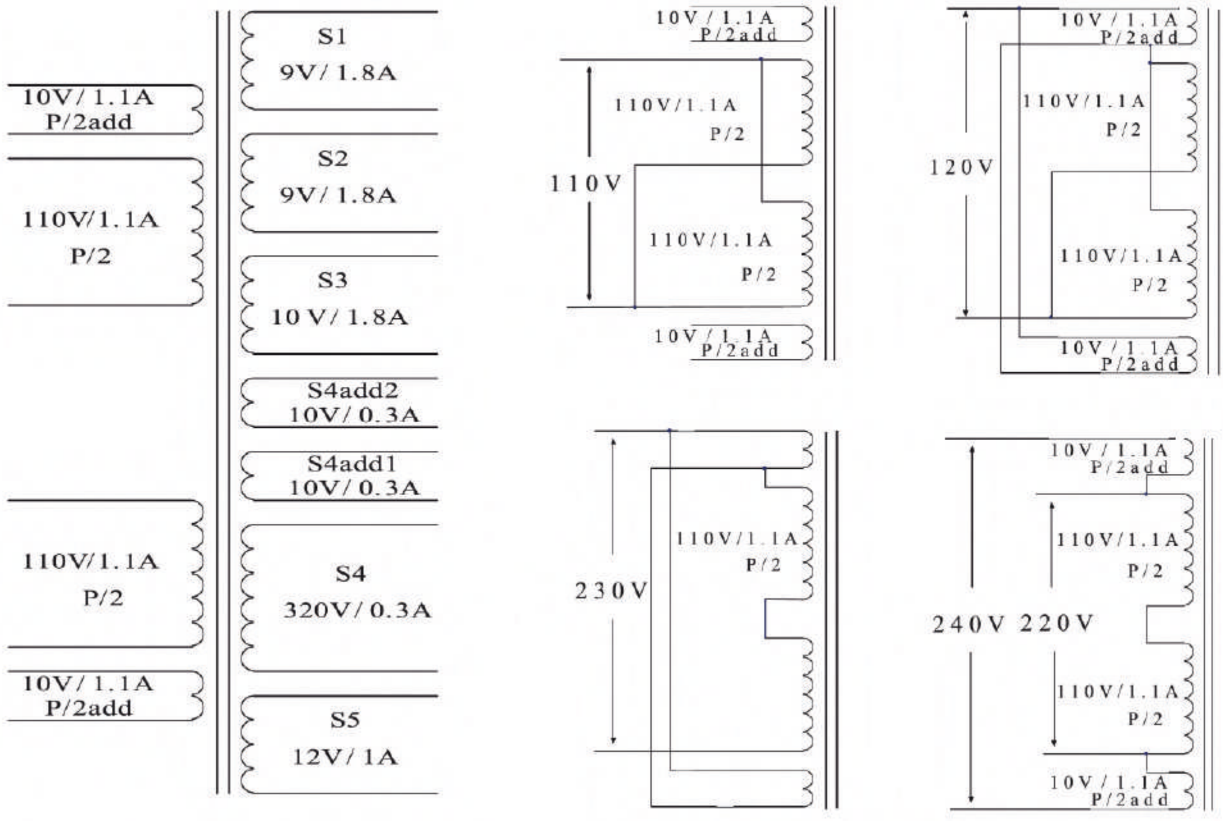
$I_{S\,[RMS]} \approx 1.5 \times 0.181A \approx$ ***272 mA***
Secondary sections of power supply transformer:

Secondary 1: S1 – 9 V / 1.8 A (300B heater)
Secondary 2: S2 – 9 V / 1.8 A (300B heater)
Secondary 3: S3 – 10 V / 1.8 A (6SN7 heater)
Secondary 4: S4 – 320 V / 0.3 A (high voltage)
Secondary 5: S5 – 12 V / 1 A (auxiliary voltage)

Secondary power:
$P_S = (U \times I)_{S1} + (U \times I)_{S2} + (U \times I)_{S3} + (U \times I)_{S4} + (U \times I)_{S5}$
$P_S = 9 \times 1.8 + 9 \times 1.8 + 10 \times 1.8 + 320 \times 0.3 + 12 \times 1 = 158.4\ W$

Taking into account the power losses of the transformer and the secondary inrush current, the power of the Primary must be about 1.5 times higher than the secondary power ($\boldsymbol{P_P} = 1.5 \times P_S = 1.5 \times 158.4 \approx$ ***238 W****).*

 # Advice from practice
Additional high voltage secondary sections: calculated secondary voltage (320 V) + 10 V section + 10 V section
Primary: two primary coils $U_{mains} / 2$, *(2 × (110 V + 10 V additional coil)) — to meet specific AC line voltage standards (110 V, 120 V, 220 V, 230 V, 240 V).*



Rectifier diodes:

V_{RRM} (Maximum continuous reverse voltage).
Bridge rectifier configuration:

$$V_{RRM} > \sqrt{2} \times U_S = 1.41 \times U_S = 1.41 \times 320 = 451\,V$$

The choice is a V_{RRM} = 1000 V diode.
$I_{F(AV)}$ (Average rectified forward current) ≥ 181.1 mA.
The safe operating current of the diode is about half the $I_{F(AV)}$.
The choice is a $I_{F(AV)}$ = 3A diode.
Fast Si rectifier diodes that meet the above requirements: BYT 56M, UF5408, (1N5408).

Several more components need to be defined:

*Primary power transformer **fuse**. A fuse is an electrical safety component – it protects electrical equipment from excessive or over current. Two types of fuses can be used:*
***slow – blow** (time delay – **T**) and **fast – blow** (fast acting – **F**). Dimensions of the fuses are standardized. (20 × 5) mm fuse is very commonly used in practice, for example.*

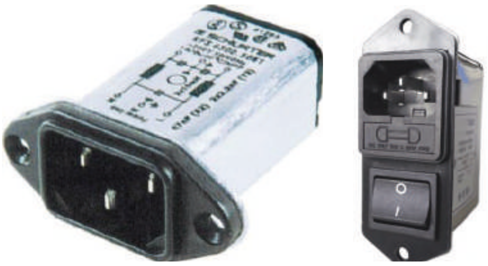
Precise calculation of the transformer primary circuit fuse is quite complicated (it is necessary to take into account complete transient process when turning on the equipment).

In practice, the rated current of the primary fuse must be ***1.35 to 2 times*** *higher than the current of the primary coil (a slow – blow fuse is used in tube amplifiers due to the high current surge at the moment of the first switching on).*
Example above: $I_{Fuse} \geq (1.35 \ldots 2) \times I_P = (1.35 \ldots 2) \times 1.1\ A \approx 1.5\ A \approx 2.2\ A;$
Standard value: ***T 2A / …V***

The rated voltage of the fuse depends on the mains voltage.
The fuse is placed in the fuse holder (fuse holder for metal chassis, fuse holder for PCB, fuse holder for wiring…).

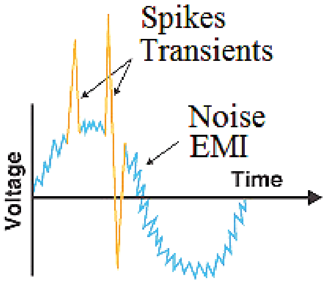
Mains switch*: 250 V / 15 A (At the moment of switching on, the primary current can be very high. Therefore, it is important to use a very robust mains switch).*

Mains filter:



One of the diseases of the modern age is a polluted mains supply. Day by day this disease is getting worse. EMI (Electromagnetic interference) filter is the first line of battle in the war against the unwanted noise and spikes generated in the mains supply line to reach and damage the audio circuit or affect the audio signal.

Types of conducting EMI (the noise travels along electrical conductors – mains wires):



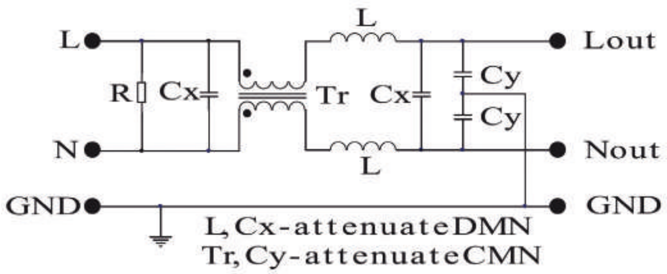
Common-mode noise (CMN) – exists on both live and neutral mains wire (they are in-phase with each other, relative to ground - the CMN current flows in the same direction on both mains wires and returns via the ground wire).
Differential – mode noise (DMN) – exists between the live and neutral mains wire (they are opposite phases – DMN current flows along live wire and returns along the neutral wire).

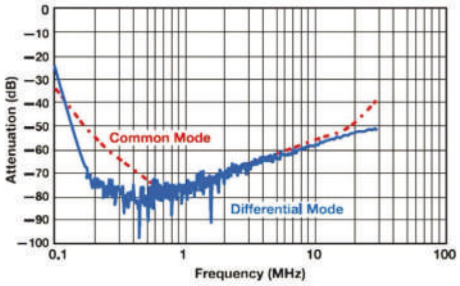
Transients: Surges and spikes – increase in voltage above the standard mains voltage (surge – voltage increase lasts more than 3ns and spike – voltage increase lasts up to 2ns).

In practice, a so-called EMI (Electro Magnetic Interference) filter is usually used in front of the Primary winding of the power supply transformer. The EMI filter reduces the magnitude of voltage spikes and high-frequency unwanted signals generated in the mains.

It is efficient in the frequency range from 150 kHz to 30 MHz. The mains socket and EMI filter can be integrated. In some designs, the fuse, mains switch and EMI filter are integrated into one common unit.
Standard rated current: 1 A, 3 A, 6 A, 10 A, …

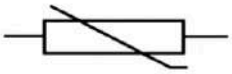
Typical EMI Filter

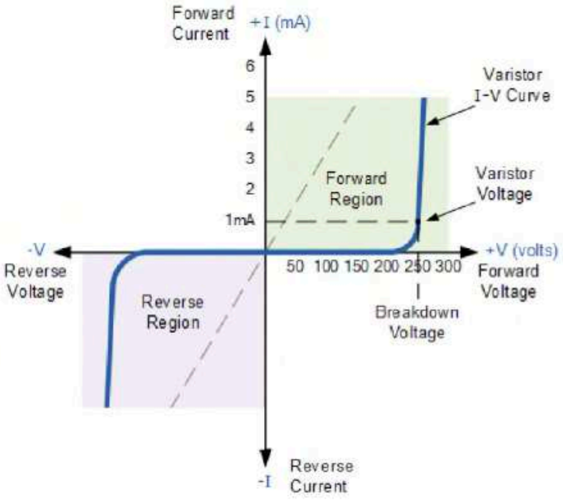




C_X – (0.1 – 1) µF class X2, C_Y – (1 – 100) nF class Y, L – (0.1 – 20) mH

Further measures to eliminate or reduce interference from the mains.
One passive component can help in the process of attenuating excessive transient voltage (spikes) from the mains:





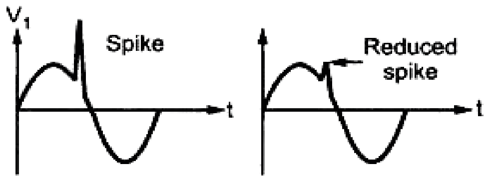
Varistor or Voltage Dependent Resistor (VDR) – resistance varies depending on the applied voltage, but in a specific way.

Current – voltage characteristic is similar to I – U characteristic of the zener diode with the difference that the varistor has symmetrical bidirectional characteristic.

As can be seen from the characteristic, the VDR resistance is high and the current flowing through the VDR is low in the region of the applied voltage lower than the breakdown voltage or the so-called clamping voltage. When the voltage (either polarity) exceeds the VDR clamping voltage its resistance decrease very fast. So, if the voltage suddenly exceeds the VDR clamping voltage (as in a case of spikes), the VDR resistance becomes extremly low and conducts the current (shunts the current created by excessive voltage). When the voltage falls below the clamping voltage, the VDR stops conducting current.

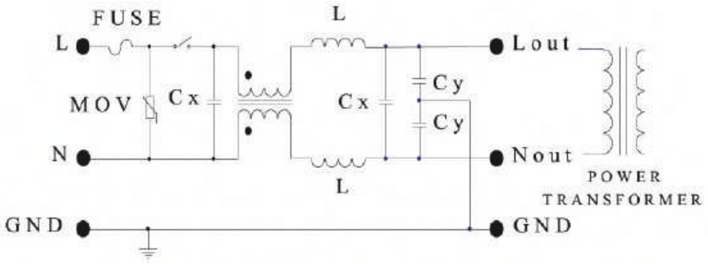
Metal oxide type of VDR or MOV (metal oxide varistor) is commonly used in practice.
In order to apply MOV in a specific circuit, it is necesseary to know the tecnical characteristics of MOV:

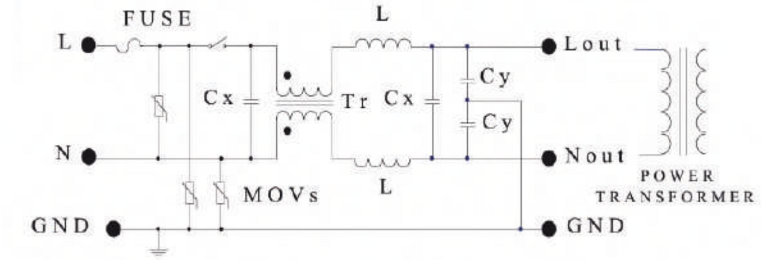
Clamping voltage (the most important) – The voltage accros the MOV when the MOV resistance becomes extremly low and the MOV starts to show significant conductivity.
(In practice MOV clamping voltage must be slightly above the mains voltage: for a mains voltage of 120V a MOV of 130 V_{RMS} to 150V_{RMS} is a good choice, for 230 V AC line power a MOV of 260 V_{RMS} to 275 V_{RMS} is a good choice).

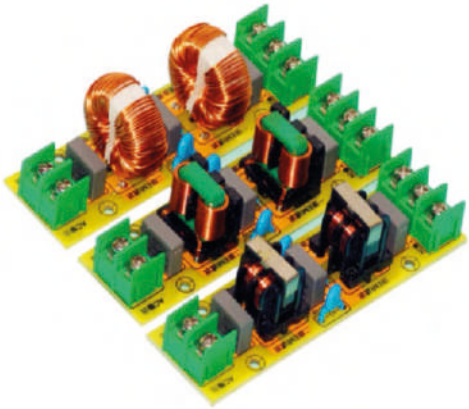


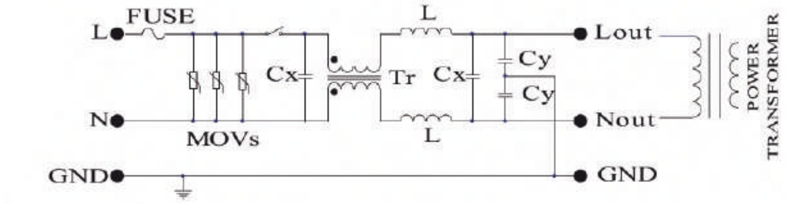
Energy absorption (dissipation) – energy (Joules) that MOV dissipates in its conductive state. 200 to 600 Joules are common values. Parallel conection of several MOVs increases the energy absorption (dissipation).

Response time – this is the time for the MOV to start conducting after applying an overvoltage pulse. Typical value: 1 to 100 nS.









*MOV can be connected between live and neutral mains wire or between three pairs of mains wires (live, neutral and ground) in front of EMI filter or behind the EMI fiter if it is compact with a mains socket, but **always behind the mains fuse.***

<u>Components pick</u>

***Tubes**:*
Use tubes from current production and carefully study the manufacturer's data sheets for each tube.

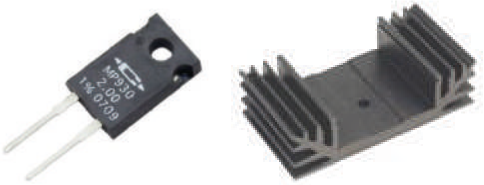
Socket:
Each tube needs a tube socket.
6SN7 is an octal base tube - octal ceramic socket for mounting on a metal chassis with mounting flange (ring – the sockets can be mounted on the top or bottom of the chasses), (or octal ceramic socket for PCB mounting).
300B: U4A ceramic socket for mounting on a metal chassis.

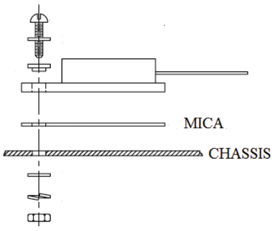
Resistors:

The use of a low noise metal film resistors is useful wherever possible, especially in the signal path.

Power resistors:

Limiting resistor R_{Lim} *applied in a delayed circuit of high voltage power supply and resistor R600 applied in a heater power supply – power wire wound resistors in ceramics.*





300B cathode power resistor – power ceramic resistor or non inductive power metal film resistor such as MP925 – 25 W, 400 Ω + 500 Ω, (must be used with heat sink. Thermal resistance of this resistor is 5 K/W and max. voltage 500 V).

The metal chassis can be used as a heat sink, and the metal film resistors can be mounted on the chassis.

Capacitors:

Capacitors applied in a signal path - ***coupling capacitors*** *(* C_S *) – Audio grade, high quality capacitors such as metal foil capacitors MKP or Teflon film PTFE or FEP and rated voltage of 400 V min. Axial leads for point – to – point wiring (radial leads for PCB).*
Capacitors connected in parallel with the diodes of high voltage bridge rectifier – metal foil capacitors MKS, rated voltage 630 V,

High voltage ***filter*** *capacitors (470 μF / 350 V): Aluminum electrolytic capacitors – high grade Al capacitors rated voltage of 350V min, rated operating temperature of 105* ^{0}*C, ESR < 0.5 Ω and ripple current (at 100 Hz) > 2 A.*
Date of production – no more than 1 year. Snap – in type for PCB mounting.
Filter capacitors used in a heater low voltage supply circuit (C601, C602) – 25 V, 85 ^{0}*C, snap – in type for PCB mounting.*

Diodes:

*Bridge rectifier diodes used in the **heater** supply circuit - low voltage, high current, low forward voltage drop, fast switching such as Schottky type 1N5822 40 V / 3 A or 1N5402 commonly used general purpose diode.*

*Bridge rectifier diodes used in the **high voltage** power supply circuit – Fast Si rectifier diodes:*
3 A / 1000 V such as BYT56M, UF5408 or 1N5408 commonly used general-purpose diodes.

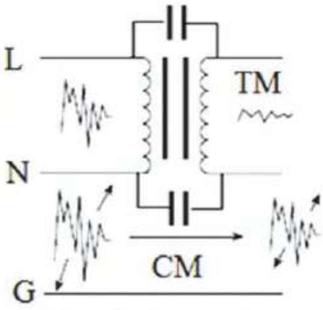
Transformers:
The electrical characteristics of power transformer are already defined above, but it is good to mention some technical aspects of the actual construction of the transformer that can affect the characteristics of audio equipment – especially hum and noise. The power transformer is the second line of battle in the war against unwanted noise and spikes that are generated in the mains so as not to penetrate the audio circuit and damage it or affect the useful signal. But its magnetic nature causes the radiation of electromagnetic energy that can reach the signal path of the amplifier and produce unwanted electromagnetic interference.

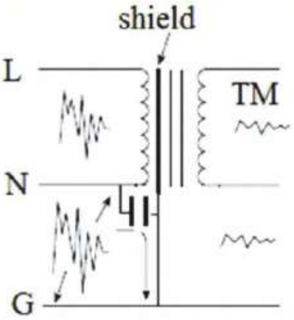
The power transformer can radiate unwanted electromagnetic energy or conduct interference signals in two ways:

- *The so-called Common Mode (CM) or line-to-ground electrical noise generated in the mains and transferred via the capacitance between the primary and secondary windings of the transformer. This noise occurs between both lines of a mains and ground. Common Mode noise is very harmful because it bypasses the power supply filters of audio equipment. The electromagnetic coupling between primary and secondary windings of the transformer is responsible for the so-called Transverse Mode (TM) noise (less harmful).*

- *Propagation of electromagnetic fields.*

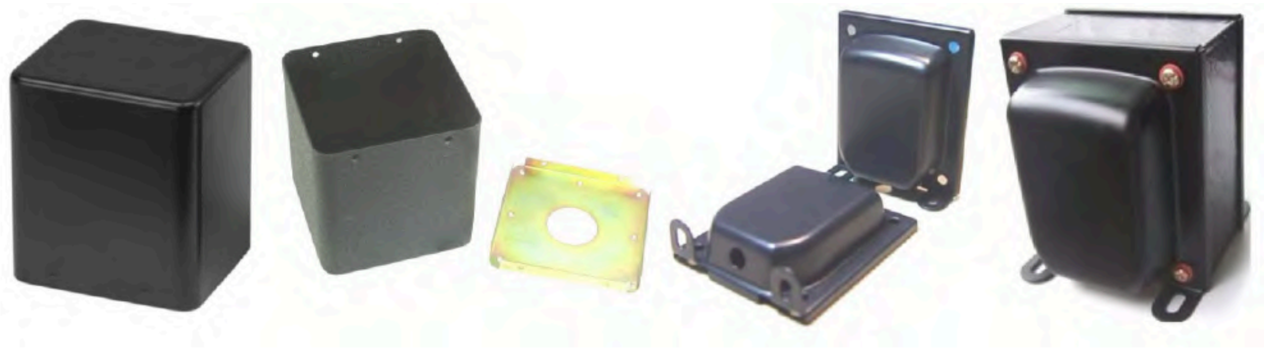
Both of these phenomena are undesirable and must be attenuated as much as possible.
*An electrostatic Faraday shield inserted between the primary and secondary windings of the transformer can lower the inter windings capacitance and attenuates CM noise. It can be made by inserting one **unclosed** and grounded turn of Cu foil wrapped around the primary windings (or by inserting one layer of Cu wire winding – one end grounded, the other end unconnected).*





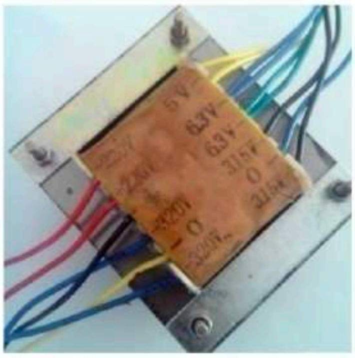
Power transformer electromagnetic radiation can be minimized by using the metal housing of the transformer as a magnetic shield.

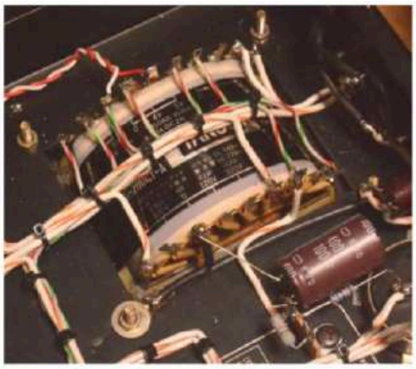
Two types of electromagnetic and magnetic field shield are used:
Two metal half covers mounted on both sides of the transformer or a metal box in which the transformer is placed.



The efficiency of the metal shield depends on the electrical and magnetic characteristics of the shield material such as electrical conductivity and magnetic properties (higher permeability of the iron sheet – higher efficiency, such as an iron - nickel alloy) and the thickness of the shield material (thicker iron sheet – higher efficiency), mainly.

It is useful that the power supply transformer is shielded. If it is completely mounted on the chassis, one of the types of shields shown above can be used. If the transformer core is mounted and bolted to the chassis and one part of the transformer passes through the hole on the chassis, one shield cover can be used:





 # Note
The electromagnetic radiation of a power transformer can be reduced by using an iron core of larger cross-section and by increasing the number of turns of the transformer windings (lower operating flux density).

As the strength of the electromagnetic field decreases in proportion to the square of the distance, the power transformer must be located as far as possible from the path of the signal and its components.
Output transformer *of a 300B single ended amplifier:* ***3500 Ω / 4 Ω / 8.5 W,*** *ready-made or made as explained in the chapter on transformers. It must be shielded.*

Choke
5H / 250mA ready-made or made as explained in the chapter on transformers. It has to be shielded too.
(10H / 250mA choke can be used for better filtration).

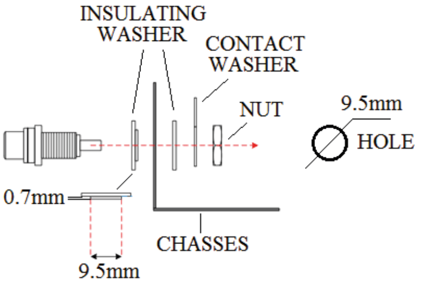
Mains inlet
Integrated mains inlet with EMI 6 A filter, chassis mounting.

Fuse holder
Chassis fuse holder for (20 × 5) mm glass fuse (T2A).

Mains switch
Any type of high quality 250 V / 15 A mains switch, chassis mounting.

Input signal connectors

The standard audio signal connector commonly used in a power amplifier and preamplifier is a Cinch (RCA or a phono connector). It is designed for chassis and PCB mounting.

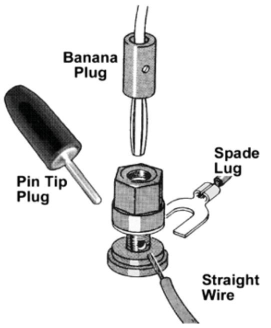


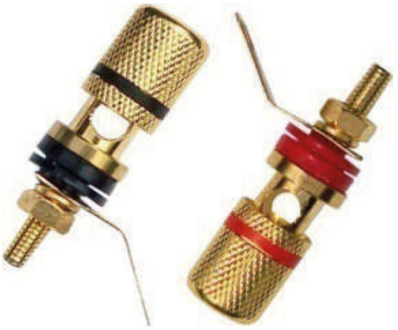
The cinch connector for mounting on a metal chassis must be insulated from the chassis and it is designed with insulating washers (carefully measure the diameter of the insulating washer before drilling holes in the chassis).

Standard cinch color:
WHITE – Left channel
RED – Right channel

One of the main characteristics of quality cinch connectors is its high galvanic conductivity (Cu body plated by gold, silver or rhodium), as well as a tight and reliable mechanical contact.

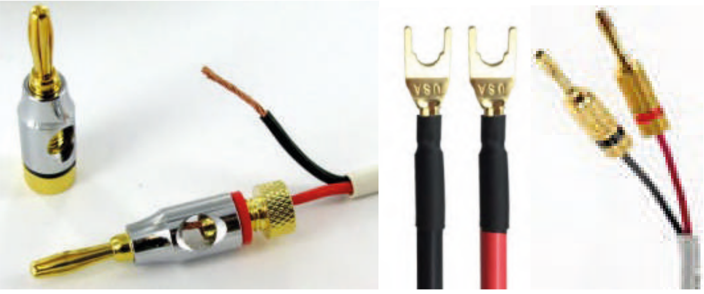
Loudspeaker connectors





The current that flows through the speaker outlet of the audio amplifier can be quite high, so the speaker connectors must be designed to handle a current of a few amperes (tight and reliable mechanical contact and good galvanic conductivity). For practical reasons the most modern loudspeaker connectors are designed to provide several types of connection modes:

banana plug, lug terminal, bare wire inserted through the connector body hole, pin connector... (so- called five – way binding post)



The speaker connectors are isolated from the metal chassis using the insulating washers. Before drilling holes in the chassis, it is necessary to carefully measure the diameter and the shape of the insulating washers.

Features of a typical Hi End speaker connectors - five way binding post:

- Current handling up to 8A
- Accepts bare wire up to AWG8 (8mm^2)
- Accepts 4mm banana plug
- Lug terminal, 6.35mm inner diameter
- Mounting hole size: 6.8mm diameter
- Standard color: red and black

In order to obtain the best possible electrical conductivity, five way binding post is made of copper or brass plated by nickel, silver, gold, rhodium,...

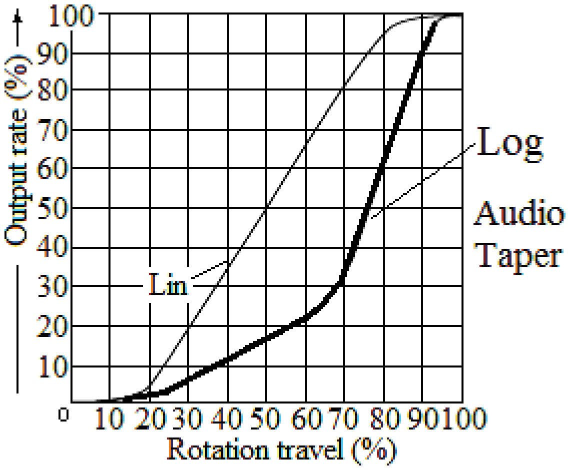
Volume potentiometer

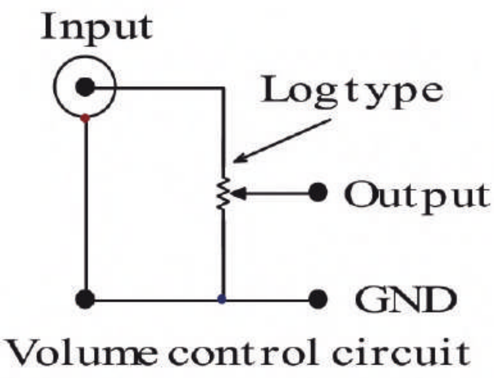
Hi End quality rotary single-turn or dual gang ***Log type*** *(commercial: "audio taper") potentiometer. There is a confusion in marking the type of potentiometers by the manufacturers:*

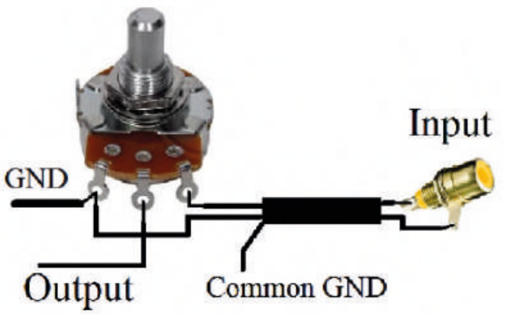
Type	String	Asia	Europe	USA
Lin	LIN	B	A	B
Log	LOG	A	C	A

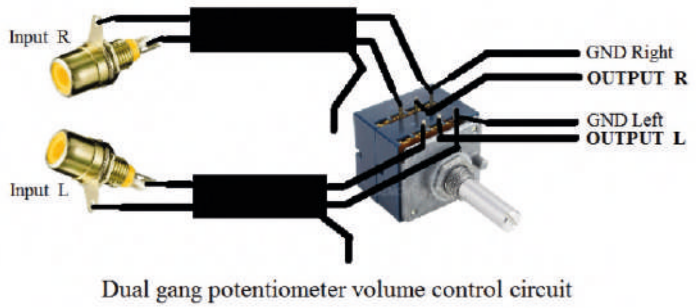
It is necessary to consult the manufacturer's datasheet to make the right decision.

Volume control circuit wiring:

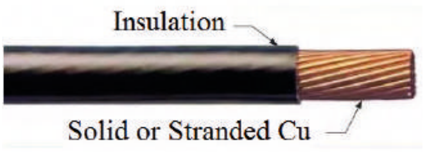








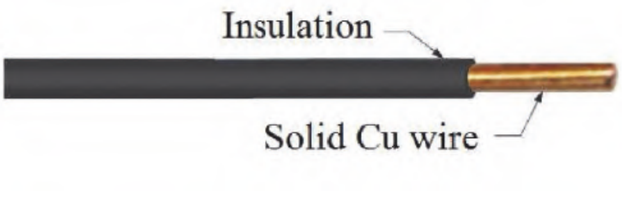
Wires

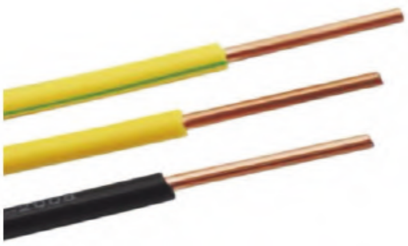


A tube amplifier chassis may carry lethal AC and DC high voltages.

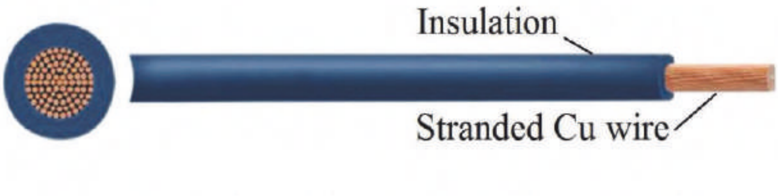
For safety reason, the basic rule is: use electrical conductors – wires with appropriate ***insulation****.*
Wires of different cross sections, insulation and electromechanical design are used for wiring inside the amplifier.
It is common to use Cu wires and their variations such as tinned, silver-plated or nickel-plated Cu wires.
Wires can be designed as ***Solid*** *and* ***Stranded.***

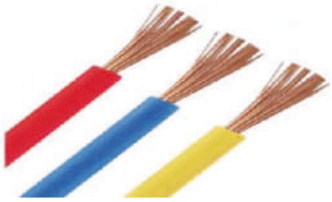
Solid wire: one solid strand





Stranded wire: *a number of solid wires of smaller cross-section twisted together to form a single, larger conductor. The main advantage of stranded wire over solid wire is its mechanical flexibility.*





There are many types of stranded wires on the market in terms of design related to the number and diameters of strands, ways of twisting…

Wire load carrying capacities

Wire current carrying capability guideline (Basic standard characteristic of wire: Americen Wire Gauge. (AWG) and Metric Gauge – size):

AWG gauge	Conductor Diameter Inches	Conductor Diameter mm	Conductor cross section in mm^2	Ohms per 1000 ft.	Ohms per km	Maximum amps for chassis wiring	Maximum amps for power transmission
0000	0.46	11.684	107	0.049	0.16072	380	302
000	0.4096	10.40384	84.9	0.0618	0.202704	328	239
00	0.3648	9.26592	67.4	0.0779	0.255512	283	190
0	0.3249	8.25246	53.5	0.0983	0.322424	245	150
1	0.2893	7.34822	42.4	0.1239	0.406392	211	119
2	0.2576	6.54304	33.6	0.1563	0.512664	181	94
3	0.2294	5.82676	26.7	0.197	0.64616	158	75
4	0.2043	5.18922	21.1	0.2485	0.81508	135	60
5	0.1819	4.62026	16.8	0.3133	1.027624	118	47
6	0.162	4.1148	13.3	0.3951	1.295928	101	37
7	0.1443	3.66522	10.6	0.4982	1.634096	89	30
8	0.1285	3.2639	8.37	0.6282	2.060496	73	24
9	0.1144	2.90576	6.63	0.7921	2.598088	64	19
10	0.1019	2.58826	5.26	0.9989	3.276392	55	15
11	0.0907	2.30378	4.17	1.26	4.1328	47	12
12	0.0808	2.05232	3.31	1.588	5.20864	41	9.3
13	0.072	1.8288	2.63	2.003	6.56984	35	7.4
14	0.0641	1.62814	2.08	2.525	8.282	32	5.9
15	0.0571	1.45034	1.65	3.184	10.44352	28	4.7
16	0.0508	1.29032	1.31	4.016	13.17248	22	3.7
17	0.0453	1.15062	1.04	5.064	16.60992	19	2.9
18	0.0403	1.02362	0.823	6.385	20.9428	16	2.3
19	0.0359	0.91186	0.653	8.051	26.40728	14	1.8
20	0.032	0.8128	0.519	10.15	33.292	11	1.5
21	0.0285	0.7239	0.412	12.8	41.984	9	1.2
22	0.0253	0.64516	0.327	16.14	52.9392	7	0.92
23	0.0226	0.57404	0.259	20.36	66.7808	4.7	0.729
24	0.0201	0.51054	0.205	25.67	84.1976	3.5	0.577
25	0.0179	0.45466	0.162	32.37	106.1736	2.7	0.457
26	0.0159	0.40386	0.128	40.81	133.8568	2.2	0.361
27	0.0142	0.36068	0.102	51.47	168.8216	1.7	0.288
28	0.0126	0.32004	0.080	64.9	212.872	1.4	0.226
29	0.0113	0.28702	0.0647	81.83	268.4024	1.2	0.182
30	0.01	0.254	0.0507	103.2	338.496	0.86	0.142
31	0.0089	0.22606	0.0401	130.1	426.728	0.7	0.113

32	0.008	0.2032	0.0324	164.1	538.248	0.53	0.091
Metric 2.0	0.00787	0.200	0.0314	169.39	555.61	0.51	0.088
33	0.0071	0.18034	0.0255	206.9	678.632	0.43	0.072
Metric 1.8	0.00709	0.180	0.0254	207.5	680.55	0.43	0.072
34	0.0063	0.16002	0.0201	260.9	855.752	0.33	0.056
Metric 1.6	0.0063	0.16002	0.0201	260.9	855.752	0.33	0.056
35	0.0056	0.14224	0.0159	329	1079.12	0.27	0.044
Metric 1.4	.00551	.140	0.0154	339	1114	0.26	0.043
36	0.005	0.127	0.0127	414.8	1360	0.21	0.035
Metric 1.25	.00492	0.125	0.0123	428.2	1404	0.20	0.034
37	0.0045	0.1143	0.0103	523.1	1715	0.17	0.0289
Metric 1.12	.00441	0.112	0.00985	533.8	1750	0.163	0.0277
38	0.004	0.1016	0.00811	659.6	2163	0.13	0.0228
Metric 1	.00394	0.1000	0.00785	670.2	2198	0.126	0.0225
39	0.0035	0.0889	0.00621	831.8	2728	0.11	0.0175
40	0.0031	0.07874	0.00487	1049	3440	0.09	0.0137

Note:
The rated current carrying capability depends on many factors: type (solid or stranded) and cross-section of the wire, electrical characteristics of wire material (electrical conductivity, thermal characteristics), thermal and mechanical characteristics of insulation, thermal conductivity of the insulation material, air convection, required or tolerated voltage drop, … Therefore, practical and approximate methods are used to choose the size of the wire.
For wiring inside the amplifier chassis, one quick and easy way to calculate the maximum current capacity of Cu wire:

$$\text{Cu Wire Capacity [A]} = (4 \div 6) \times \text{Size of wire (cross section) [mm}^2]$$

Example:
For 1-mm² cross-section wire, the wire capacity is $(4 \div 6) \times 1\ mm^2 = (4 \div 6)\ A$
Minimum wire size (minimum cross-section):
Size of wire (cross-section) [mm^2] = Required wire current handling [A] / $(4 \div 6)$

Example:
Required minimum wire size (cross section) for wiring a circuit with a load current of 2 A:
$2\ A\ /\ (4 \div 6) = (0.33 \div 0.5)\ mm^2$

For wiring the tube heater power supply it is better to use: Size of wire (cross-section) [mm^2] = I_f [A] / $(3 \div 4)$

Wire insulation
Inside the tube amplifier chassis, the temperature can be very high (it is good to keep temperature below 60^0C by appropriate chassis design), the wires can be located near components with high thermal dissipation and high surface temperature, used insulated wires must operate in such ambience. Rated temperature of insulation – the maximum temperature at which the insulated wire can operate without damaging the insulation.

The insulating materials commonly used to cover Cu wire:
Insulating material based on plastic materials.

- ***Thermoplastic (****PVC – Polyvinylchloride, PE – Polyethylene, ECTFE – Ethylene Chlorotrifluorethylene, PVDF – Polyvinylidene Fluoride, Nylon,…).*
  Thermoplastic insulation softens and can even melt when exposed to a sufficiently high temperature.
  Rating temperature is usually in the range of -55^0C to $(60^0C \div 105^0C)$.
  The low price of this type of insulated wire is the main reason for its mass use.

- ***Thermoset***
  Thermoset insulation does not soften when exposed to high temperature.(based on EPR – Ethylene Propylene and PTFE – Polytetrafluoroethylene). The most widely used thermoset insulation material is the well- known Teflon.
  Rated temperature is usually up to 200^{0}C .

- ***-Rubber insulation***
  The most widely used is the well-known silicone rubber.
  Rated temperature is usually up to 180^{0}C .

Wire coding

The type of plastic insulation can be identified by a code:

- T – Thermoplastic
- H – Heat-resistance (HH – higher heat resistance)
- N – Nylon jacket
- W – Wet location

Example:
THHN – thermoplastic, higher heat resistance and nylon jacket.

Rated voltage of insulated wire

Wires inside the chassis can be located near the components or other wires at the high voltage. Also, the wires at high voltage can be located near the grounded walls of the chassis.

The rated voltage depends on the type of the insulation material used (electrical characteristics) and its thickness.

Usually, the rated voltage (maximum voltage to which it can be connected) is expressed by two values of alternating current U_0 / U *[V]:*

U_0 – *voltage between conductor and earth (ground).*
U – *voltage between conductors.*

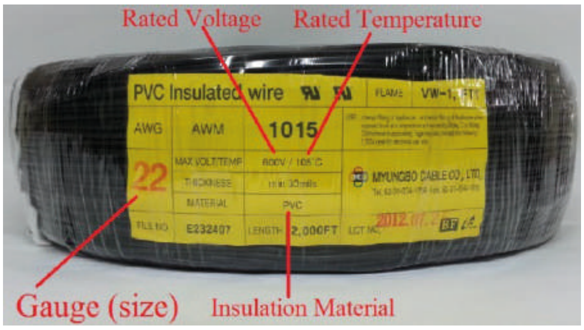
Example:
600 / 1000V: insulated wire capable of withstanding a voltage of 600 V_{RMS} *between conductor and earth (ground) and 1000* V_{RMS} *between adjacent conductors.*

In the technical data of the manufacturer, it is not unusual to find only the rated voltage of the wire insulation, such as 300 V, 600 V, 1000V, and it refers to the maximum operating voltage.

Technical specification of insulated wire that needs to be know in order to make the right choice of wire for a specific purpose:

- Size (Gauge), cross section or diameter
- Current load capacity
- Rated voltage of insulated wire
- Rated temperature of insulated wire
- Type of insulation (thermoplastic – such as PVC, thermoset – such as PTFE (Teflon), rubber – such as Silicone).

The technical specification of the wire can usually be found on the reel or drum of the wire.



The choice of wire (example)

Heater wiring
Solid or stranded insulated Cu wire.
6SN7: 6.3 V / ***0.6 A*** *(one channel)*
Minimum wire cross-section [mm^2]:
Current handling [A] / (3 ÷ 4) = 0.6 [A] / (4 ÷ 6) = ***(0.2 ÷ 0.15) mm^2***
Standard wire gauge table, nearest wire size: AWG 24 (0.205 mm^2) and AWG 25 (0.162 mm^2)
The size of the solid insulated wire (diameter - d [mm]) corresponds to the cross section AWG 25 (0.162 mm^2):

Cross section = $(d/2)^2 \times \pi \rightarrow d[mm] = 2 \times \sqrt{\frac{Cross\ section\ [mm^2]}{\pi}} = 2 \times \sqrt{\frac{0.162}{3.14}} = 0.454\ mm$,

nearest metric standard wire: d = 0.5 mm.

Rated voltage of insulated wires: The wires used for wiring tube heaters can be located near components or uninsulated wires at a high voltage that is over 440 V_{DC} in the example above. Therefore, the rated voltage of insulated wire must be: 600V, 600/1000V.

Rated temperature of insulation: > 90^{0}C. Thus, the insulating material of the wire can be quality PVC or PTFE (Teflon) or silicone rubber.

Wire for wiring 6SN7 heater:

- Solid or stranded Cu wire.
- Size: Stranded AWG 25 (0.162 mm^2) to AWG 22 (0.327 mm^2), metric 0.25 mm^2 can be used or solid wire: d = 0.5 mm to 0.7 mm.
- Insulation: PVC, PTFE (Teflon), Silicone rubber.
- Rated voltage: 600/1000 V, 600 V.
- Rated temperature: 105 ^{0}C.

 # Note
Each tube heater is wired with a twisted pair of wires.
If the heaters of two tube are wired in parallel from the same heater power supply source, the wire size must be recalculated to handle twice the current required to power heater of one tube.
Parallel connected heaters two 6SN7tubes: 2 × 0.6 A = 1.2 A.

Wiring between the secondary of the power transformer and the common DC rectifier and the stabilization circuit for heater power supply two 6SN7. Wire current handling is: 2 × 0.6 A = 1.2 A.
Insulated stranded wire: AWG 18 or metric (0.75 ... 1) mm^2.

Wire for wiring 300B heater (1.3 A):

- *Stranded Cu wire*
- *Size: Stranded AWG 20 (0.519 mm^2) to AWG 18 (0.823 mm^2), metric: 0.75 mm^2*
- *Insulation: PVC, PTFE (Teflon), Silicone rubber*
- *Rated voltage: 600/1000 V, 600 V*
- *Rated temperature: 105 ^{0}C*

 # Note:
The wire of the above size is used for wiring the heater of one 300B tube.

High-voltage circuit wiring
One common high voltage power supply circuit is used to power both amplifier channels. The currents of the anode circuits of the tubes are not so high and the high voltage circuits can be wired with wires of not so large cross section. In the example above the high voltage power supply current is less than 200 mA. In practice, wires with larger cross-section than calculated one are used: insulated stranded Cu wire AWG 22 (23) or AWG 20, metric 0.25 mm^2 , 0.5 mm^2, 0.75 mm^2. Insulation: PVC, PTFE (Teflon), Silicone rubber.
Rated voltage: 600/1000 V, 600 V. Rated temperature: 105 ^{0}C.

Ground buss

The ground buss (ground bar) is used in the construction of the power amplifiers for two reasons: to conduct electrical current and as a mechanical carrier for electronic components.
Therefore, it must fulfill two main technical requirements:

- *Low electrical resistance and current carrying capability.*
- *Good soldering ability*
- *Good mechanical strength*

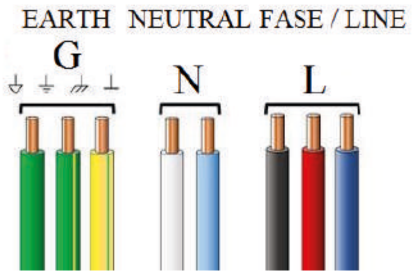
In practice, the uninsulated solid Cu wire is most commonly used. Because it must handle a current equal to the sum of the supply currents of all amplifier stages (input, driver, output, and even the secondary load current of the output transformer), its cross section must be large to provide low resistance and therefore minimal voltage drop along the bus.

Cu (it is good to be oxygen free Cu - OFC) uninsulated wire AWG 14 (2.08 mm^2) or AWG 12 (3.31 mm^2), metric d = 1.6 mm to 2 mm is used as ground buss usually in medium power amplifiers.

Wires used for wiring electronic components of mains supply circuit (AC circuits) .

The wiring of the mains socket (or EMI filter with built-in mains socket, mains fuse and power switch) and the winding of the primary power transformer must be made using high quality insulated (PVC, PTFE – Teflon, Silicone; rated voltage of 600/100V; Temperature of 105^0C) stranded Cu wires of appropriate cross section (AWG 18 to AWG 14, metric 1mm^2 to 2mm^2).

Color Code:

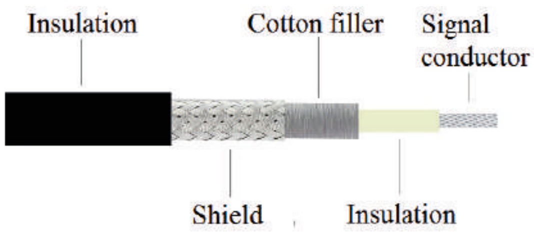


GROUND (EARTH):	*Green* *Green with yellow strip* *Yellow with green strip*
NEUTRAL:	*White* *Light blue*
FASE / LINE:	*Black* *Red* *Blue*

Audio cable

Electric signals of the audio frequency range that we encounter in practice can be very low levels (MC and MM turntable cartridges - 5 μV to few mV, preamplifier output of several hundred mV). It is very difficult to transfer such signals by wire and keep original sound image.

The basic function of an audio cable is to transfer an electrical signal of the audio frequency band with the least loss and distortion and to protect the signal from the effects of unwanted noise such as RFI and EMI.



Such technical requirements can be met by using specially designed cables for use in audio equipment: an insulated signal conductor (solid or stranded) covered with braided Cu strands that act as an electrostatic shield (usually grounded).

The effectiveness of protection against EMI and RFI depends on the design and the quality of the shield – in general, the large the coverage, the better the protection.

The percentage of coverage depends on the number of braid strands, the angle of braiding, the folding number (coverage density ranges from 50% to 97%). Some cables have an additional shield made of aluminum (or Cu) foil inserted between the signal conductor and the braided Cu shield (improved shield density up to98%).

Good electrical conductivity of the shield is very important too (it depends on the electrical characteristics of the shield material).

Good conductivity and appropriate cross section of signal conductor (solid or stranded) is necessary. On the market there are a large number of audio shielded cables in various technical designs (shielded cable with one, two, three, four or even more signal conductor with different number of strands and its diameter, shield density, insulation material and fillers…).

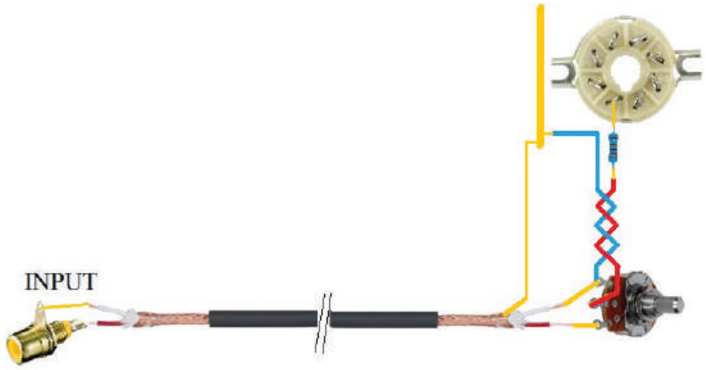
Typical characteristics of an audio cable:

- *Shielded cable with one signal conductor,*
- *Shielding: 32 strands with a diameter of 0.12 mm each (Specification: **N × n / Ø**; N is the number of signal conductor, n is the number of strands and Ø is the strand diameter [mm] - 1 × 32 / 0.12),*
- *Signal conductor: 10 strands with a diameter of 0.12 mm each, with a total cross-section of 0.11 mm^2 ,*
- *Insulation: PVC,*
- *Outside diameter: 3.7 mm.*

Shielded cable with two signal conductors

- *Signal conductor: 2 × 16 / 0.1 mm (cross section of 0.12 mm^2, each),*
  DC resistance: 158 mΩ / m, PVC insulation, outside diameter: 5 mm,
- *Signal conductor: 2 × 25 / 0.1 mm (cross-section of 0.20 mm^2 , each), DC resistance: 92.3 mΩ / m, PVC insulation, outside diameter: 5.5 mm.*

Example:
Wiring the RCA input connector and volume potentiometer



 # Note:
If the distance between the two connection points of the audio signal path is not long (less than a few cm), the wiring can be made by a pair of twisted insulated wires of appropriate cross-section.

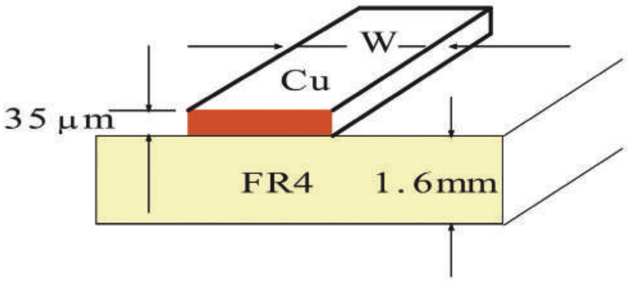
However, the longer the cable, the higher the probability of causing unwanted effects on audio signals in the form of noise and hum (buzzing) caused by RFI and EMI.

<u>Keep the signal path as short as possible.</u>
Minimizing the length of wiring can be achieved by carefully designing the layout of the electronic components on the chassis.

Tube heaters power supply

DC regulated power supply – two 5 V/1.2 A (output tubes) and two voltage regulators of 6.3 V/1.2 A (driver and input tubes).

They can be made using PCB technology as compact modules.



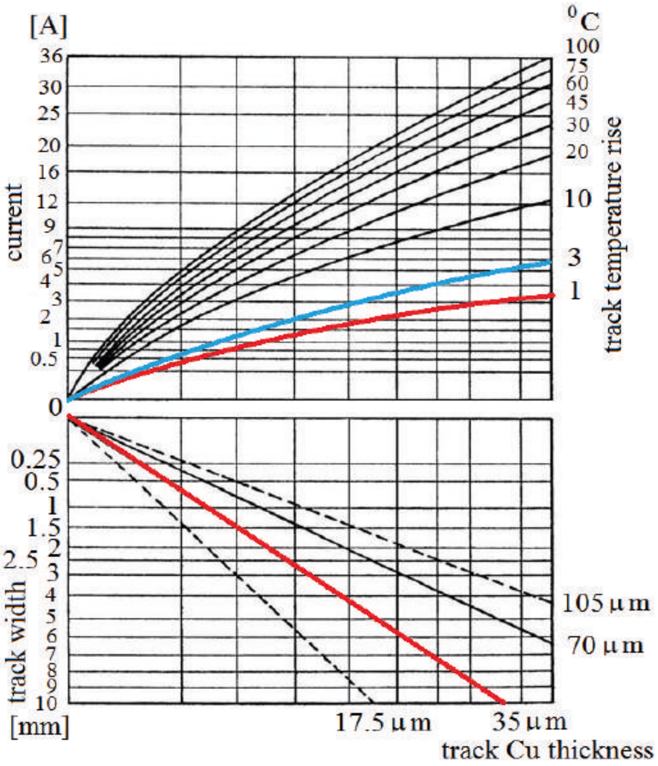
 # A few notes about PCB design (or some PCB design rules)

Common printed circuit board (PCB) material:
FR4**,** board thickness 1.6mm, one or two layers of Cu.
Cu layer thickness: 35µm
The Cu pattern can be tinned to improve the current carrying capability of the Cu traces.

The width of the **Cu trace** depends on the current flowing through the trace and the allowed increase in the temperature of the Cu trace.

The minimum trace width is published in the IPC – 2221 recommendation.
An easy way to define minimum trace width (W) is to use a diagram (or table):

Trace continuous current capacity >10sec [A]	**1^{0}C temperature increase**		3^{0}C temperature increase		10^{0}C temperature increase	
	mm	mil	mm	mil	mm	mil
0.1	0.11	4.25	0.05	2.2	0.03	1.18
0.2	0.27	10	0.14	5.5	0.07	2.60
0.3	0.46	18	0.24	9.3	0.12	4.50
0.4	0.67	26	0.35	13.5	0.17	6.50
0.5	0.90	35	0.46	18	0.23	8.80
0.6	1.15	45	0.59	23	0.28	11
0.7	1.40	55	0.72	28	0.35	14
0.8	1.67	68	0.87	34	0.42	16
0.9	1.95	76	1.00	39	0.49	19
1.0	2.24	88	1.15	45	0.56	22
1.1	2.53	100	1.30	51	0.63	25
1.2	2.84	113	1.46	58	0.71	28
1.3	3.15	124	1.65	64	0.78	31
1.4	3.48	137	1.80	70	0.87	34
1.5	3.81	150	1.96	77	0.95	37
1.6	4.15	163	2.15	84	1.03	41
1.7	4.50	177	2.30	91	1.12	44
1.8	4.84	190	2.50	98	1.20	47
1.9	5.20	205	2.70	105	1.30	51
2.0	5.56	219	2.86	113	1.38	54
2.5	7.50	294	3.85	152	1.85	73
3.0	9.50	373	4.90	192	2.35	93



The Cu pattern can be tinned to improve the current handling capability of the Cu traces.

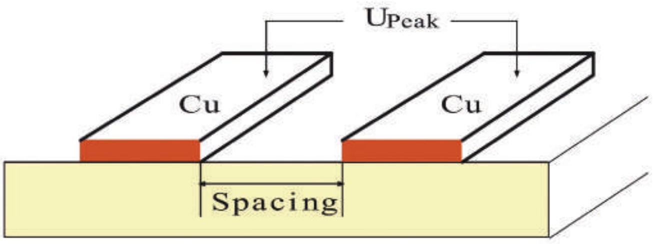
Example:
Current – 1 A
Using the table: The trace width must be 2.24 mm for a temperature rise of 1^0*C or 88 mils (1 mil = 0.0254 mm).*

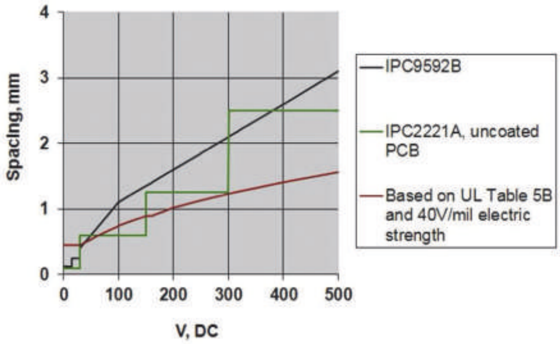
Current – 300 mA
Using the table: The trace width must be 0.46mm for a temperature rise of 1 ^{0}*C or 18 mils (1 mil = 0.0254 mm).*

Spacing – minimum distance between the two nearest traces. It depends on the voltage difference of the two nearest traces.
Spacing between traces can be calculated as:

$$\mathbf{Spacing}\ [\mathrm{mm}] = \mathbf{0.6 + 0.005 \times U_{Peak}}\ [\mathrm{V}]$$

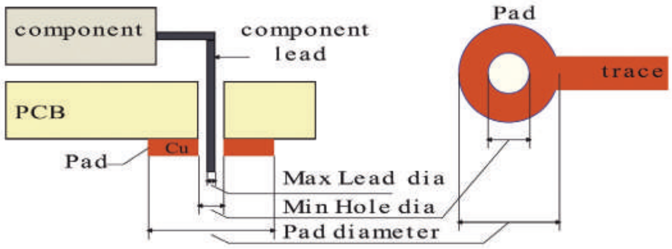
Some recommendations:





Example:
The two traces of the PCB used in the high voltage regulator circuit with a voltage difference of 400 V are next to each other.
Spacing ≥ 0.6 + 0.005 × 400 = 2.6 mm

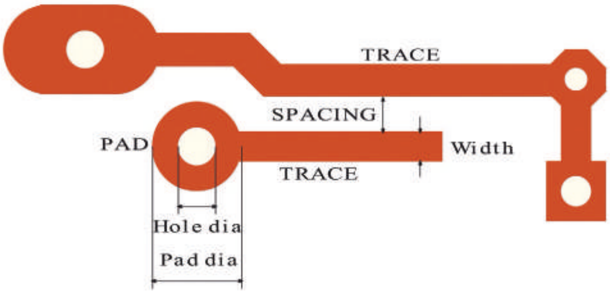
Pad diameter and hole size



Min. Hole Size = Max. Lead Dia.+ 0.25mm
Min. Pad Dia. = Min. Hole Size + 0.7mm

The diameter of the pad must be larger (or the same size) than the width of the joined trace.

The shapes of the pads can be different.



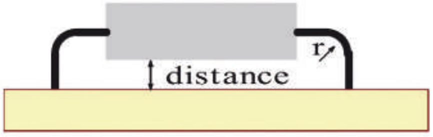
Example:
Example:
1W resistor, resistor leads diameter = 0.8 mm
Minimum hole size = 0.8 mm + 0.25 mm = 1.05 mm

The holes can be drilled using a ø1 mm drill bit.
Minimum pad diameter = Minimum hole size
+ 0.7 mm = 1 mm + 0.7 mm = 1.7 mm (67 mils).

2 – layer (double layers) PCB – Cu on a top and the bottom side of the PCB. Advantage: easy connection of pads and the traces on one side to pads and the traces on the other side of the PCB. The density of the components can be higher (number of components per square cm of PCB). The advantage is greater freedom for PCB designers to create component and trace layouts – especially if PCB holes are metalized.

Universal readymade raster PCBs and wire jumpers or wires that simulate Cu traces can simulate PCBs or can be used at an early phase of device development when numerous and quick changes in the layout and connections of electronic components are required.

The components should be placed at a certain distance from the PCB to allow better airflow around the components and better cooling. The transfer of PCB vibrations to components is lower. Component leads should not be bended at right angle – they should be bended in a radius.

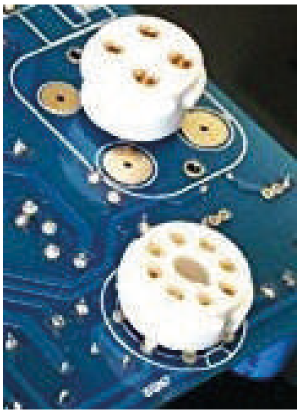


The trace carrying a high current signal (or a high amplitude voltage signal) should not be placed close to or parallel to a trace carrying a low current signal (or a low amplitude voltage signal).

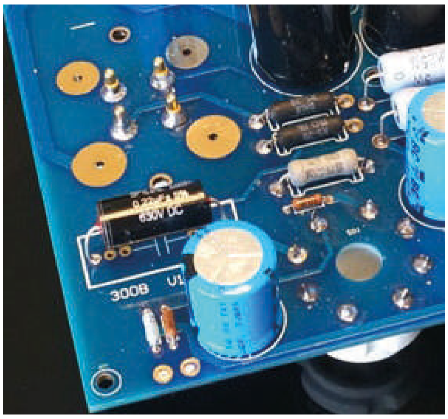
If possible (for example using a double-layer PCB), the sockets of the vacuum tubes (and tubes) should be placed on the opposite side of the PCB from the side on which the other components are placed to avoid heating the components caused by the vacuum tubes.

All components on one side of the PCB (Top side)

Tubes sockets on one side of the PCB (Cu side)



All components (except tube) on one side of the PCB (Cu side)



Example above:
Almost all parts (electronic and electromechanical components, wires,…) necessary for building amplifier, their mechanical properties, shapes and dimensions (electrical characteristics, too), are specified and defined and they must be placed and mounted on a carrier or mechanical structure where they should be wired and connected to make one functional unit – an audio amplifier.

Chassis
There are many different types of electronic components used to build audio amplifiers and each has its own specific mechanical and electrical characteristics. Some of them can be very heavy and large in size, some of them dissipate a lot of energy in normal operation and their temperature can be higher than the ambient temperature, some of them can be very sensitive to the effects of the electromagnetic interference, …

Therefore, the chassis of an audio amplifier must be designed to meet all the mechanical and electrical requirements necessary for the installation, carrying, mechanical and thermal stability and integration of all electronic and electromechanical components into one common unit, as well as to ensure their safe operation within their rated characteristics (external beauty is also important). This means that the chassis must be a mechanically stable structure capable of carrying heavy electrical and electromechanical parts, to be thermally stable structure as well as to provide good heat dissipation and air circulation / ventilation.

Materials used to build the chassis:

Different types of metal sheets:
- Cold and hot rolled steel (most common), stainless steel
- Non-magnetic materials: sheets of cooper, brass, aluminum (aluminum alloy) or aluminum profiles

Unconventional specific (esoteric) materials:
- Wood, granite, marble, onyx, alumina ceramics,…

Advantages of non-magnetic materials used to build the chassis such as cooper and aluminum sheets:
- *Not susceptible to the action of magnetic fields produced by some electronic parts such as power supply transformers for example. The transfer of the magnetic field via the chassis and its effects to the other electronic parts is minimal.*
- *Excellent electric characteristics – low electric resistance*
- *Very good thermal conductivity*
- *Corrosion resistant*
- *Aesthetics: Cu chassis can be painted with transparent lacquer, Al chassis can be painted or anodized*
- *Easy mechanical processing due to low hardness of the material (drilling, cutting, bending…).*
- *Large assortment of standard profiles on the market*
- *Acoustic properties: In general, all metals have good sound conductivity, so that side effects can occur due to poor absorption of mechanical vibrations of the chassis made of metal sheet, i.e. the chassis can be "microphonic".*

Disadvantages:
- *Low mechanical strength. The ability to carry heavy electronics parts such as transformers and chokes is low, so it is necessary to use Al sheets of greater thickness (thickness of 3mm and thicker, even 10mm thick).*
- *High price.*

Cold and rolled steel
- *Most used in practice*
- *Low cost*
- *Relatively easy for mechanical processing such as drilling, cutting, bending, welding, …*
- *Susceptible to rusting so final protection against corrosion is necessary, such as painting with thermostable paint, nickel plating, chrome plating, zinc plating , …*
- *Significantly worse electrical conductivity than Cu or Al sheet.*
- *High magnetic permeability. Unwanted feature. Prevention of this side effect is usually achieved by inserting an air gap in the magnetic field path (mounting the transformer at a distance from the chassis or inserting some non-magnetic material between the transformer and the chassis, for example), or by using a transformer shield. In practic*e, metal shielding boxes inside the chassis are also used.

Stainless steel (chrome and nickel alloy steel)

- *Non-magnetic material.*
- *Non-corrosive material (no need for additional final protection against atmospheric aggressive effects)*
- *Material of high strength and hardness.*
- *Medium costs.*
- *Can be welded without a visually noticeable welding area (after polishing)*
- *The surfaces of some types of stainless steel sheets that can be found on the market are extremely polished like mirrors.*

Disadvantages:

- *Hard to mechanical processing due to high mechanical strength and hardness of stainless steel.*
- *Medium to high prices of materials and medium to high cost of machining.*

 # Note:
Stainless steel is the best choice for building chassis, but in practice, cold and hot rolled steel sheet is most often used.

Chassis design starts with designing the layout of electronic parts. Some basic rules that must be followed when designing the layout of electronic and electromechanical parts:

- **Minimum distance between two tubes** recommended by the tube manufacturer (rated temperature of the tube glass balloon must not be exceeded even under conditions of free air circulation and some precautions must be observed). Some tube manufacturer recommendations for the distance between two tubes:
  - Small signal tubes: (1.5 ... 1.75) inch; (3.8 ... 4.5) cm
  - Medium power tubes: 2 inch; 5 cm
  - Power tubes: (2.5 ... 4) inch; (6.5 ... 10) cm;

**** Consult the manufacturer's data sheet and recommendations for specific type of tube.***

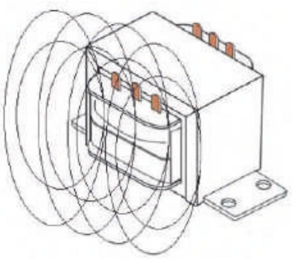
Example:
Two-channel amplifier (2 × 6SN7 and 2 × 300B) – distance between two 300B: 4 inch or 10 cm; distance between two 6SN7: more than 1.75 inch or 4.5 cm; distance between 6SN7 and 300B: more than 2 inch or 5 cm).

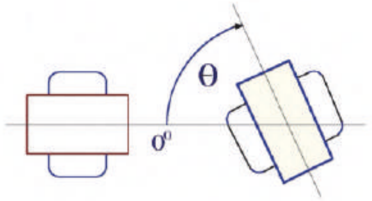
- Transformer orientation

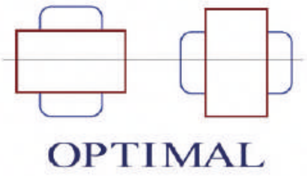
  The effect of a magnetic field originating from a source (transformer, in a case of audio amplifier: power supply transformer) on electronic parts in its environment can be analyzed by applying the well-known law of physics: magnetic interference and magnetic flux (Φ) as a measure of total magnetic field (B) passing through area (A) depends on the angle (θ) between the lines of the magnetic field and the surface (A):

  $\Phi = B \times A \times \cos(\theta)$.

  The maximum magnetic flux is for $\theta = 0^0$, $\cos(0^0) = 1$, and the minimum for $\theta = 90^0$, $\cos(90^0) = 0$ i.e. $\Phi = 0$.



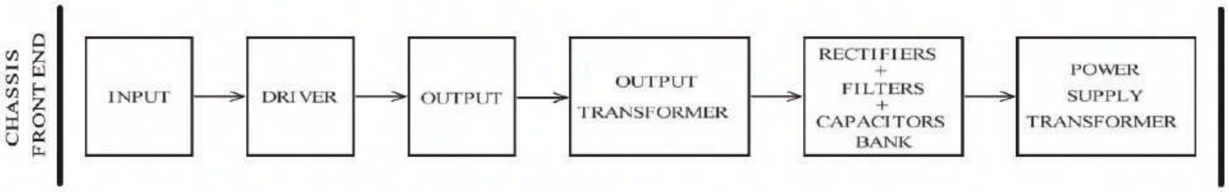




- As the strength of the electromagnetic field decreases in proportion to the square of the distance, the power transformer must be located away from the signal path of the audio amplifier and its parts.

- All wires and electronic parts conducting the audio signal must be located away from sources of unwanted electromagnetic radiation such as the power supply circuit (power supply transformers, rectifiers and filters components) and away from the mains wires inside the chassis.

- Layout of amplifier stages. One of the possible scenarios (in line):

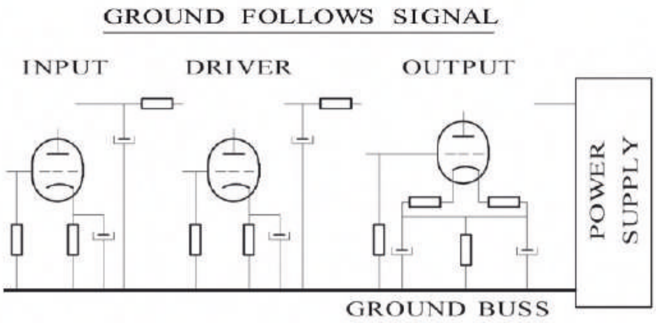


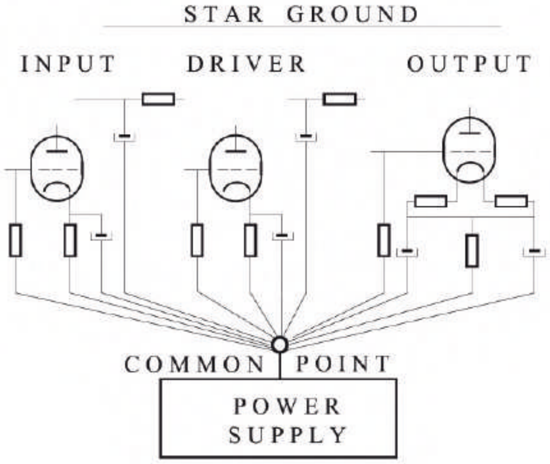
  Input stage → driver stage → output stage → output transformer → power supply rectifiers and filters → power supply transformer

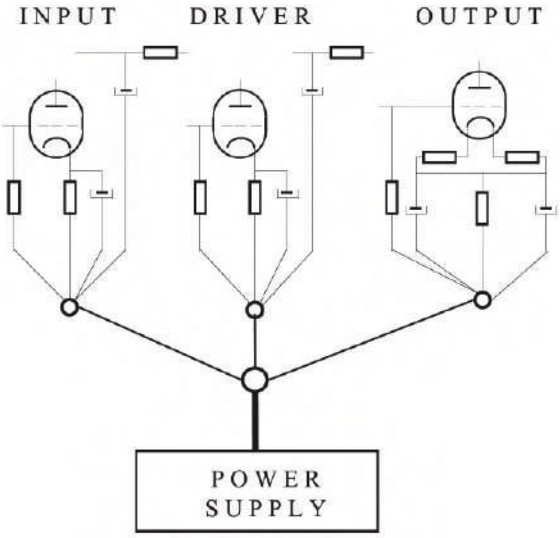
- Grounding

  Two main grounding methodologies:
    - Grounding follows the signal
    - Star-shaped grounding

  # Note: Mixed methodology is also used in practice.



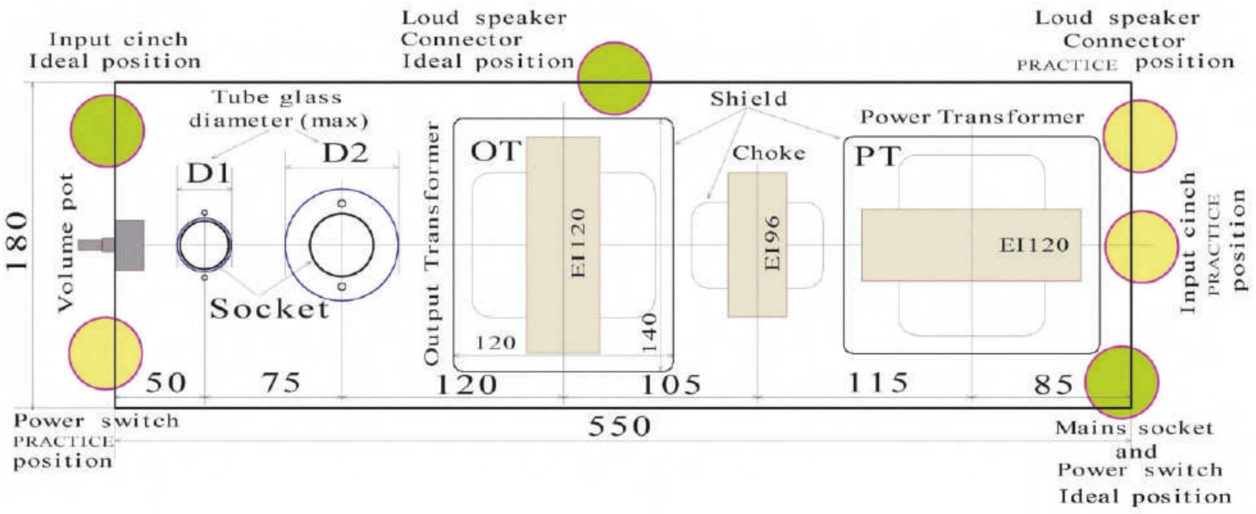




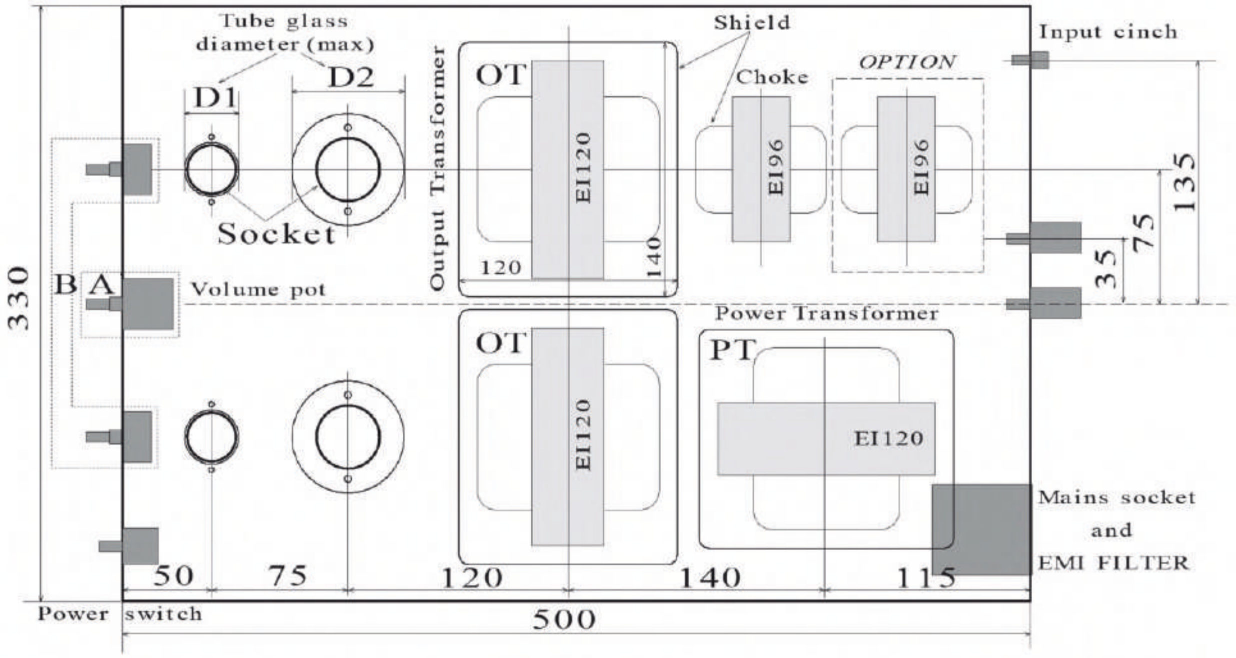
No matter which grounding methodology is used, it is important to ensure low electrical resistance and minimal voltage drop along the grounding conductors by using a wire of well-defined cross-section – high wire load carrying capacities.

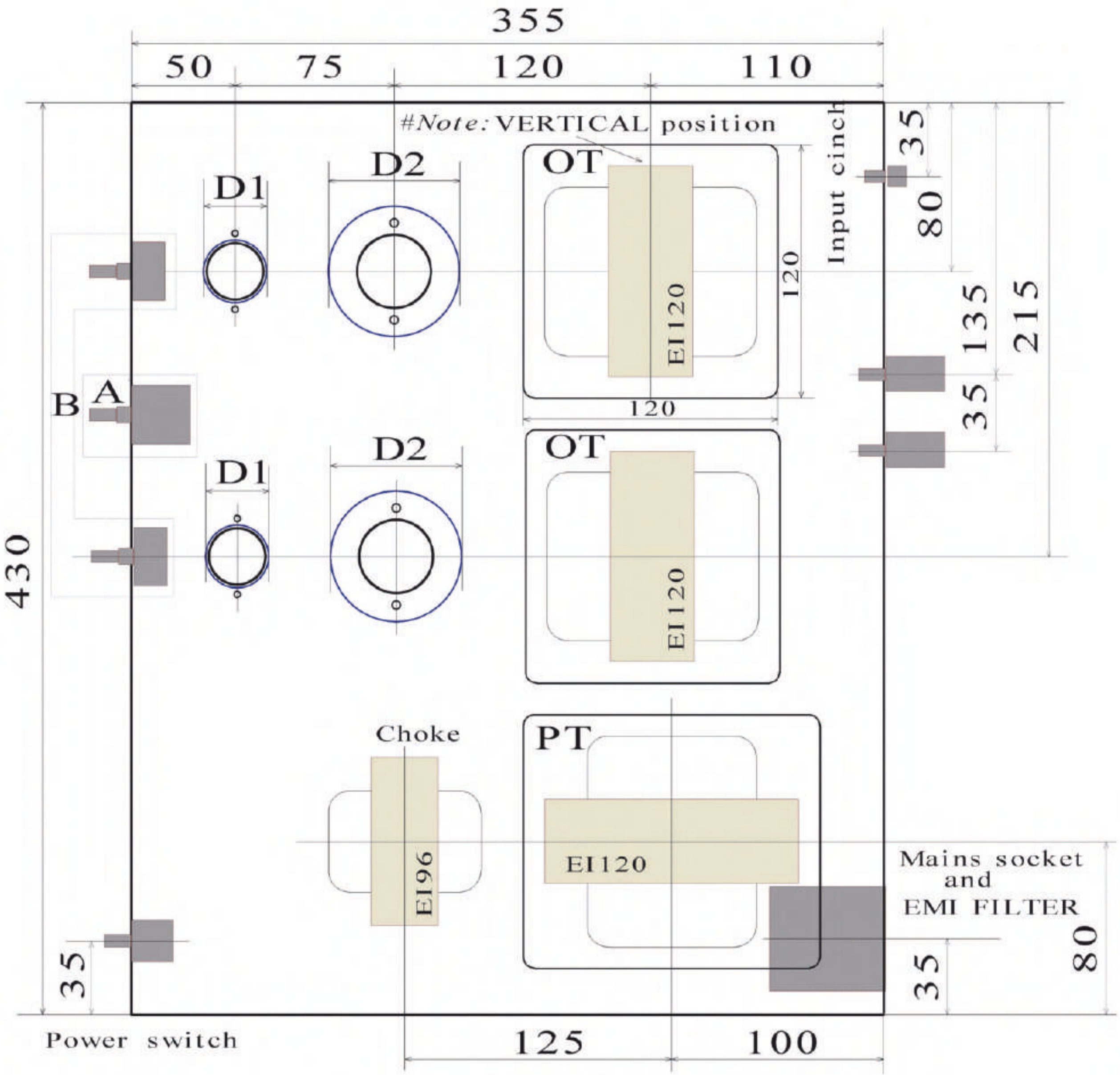
After the completion of the project phases of synthesizing the electrical scheme and defining all electronic and electromechanical components and selecting them, i.e. when their electrical and mechanical characteristics are known, including their physical dimensions, the first layout of parts (chassis layout) can be sketched, following the rules and recommendations for layout of amplifier stages, parts orientation, grounding,…

Example: Monoblock; Chassis TOP VIEW:

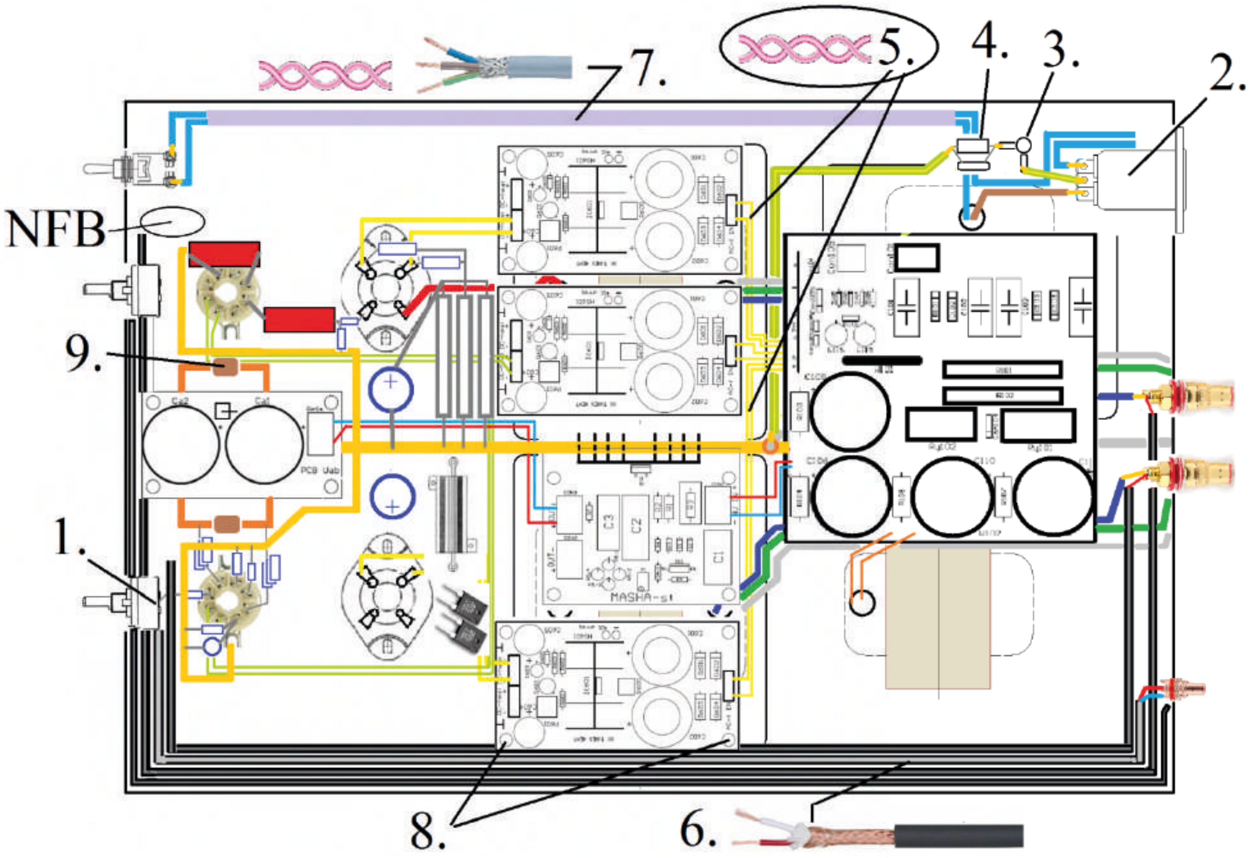


Example: Stereo Block TOP VIEW

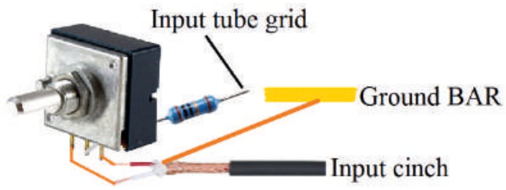



355
50
75
120
110
#Note: VERTICAL position
D1
D2
OT
EI120
120
120
Input cinch
35
80
135
215
35
B
A
D1
D2
OT
EI120
430
Choke
EI96
PT
EI120
Mains socket
and
EMI FILTER
35
80
35
Power switch
125
100

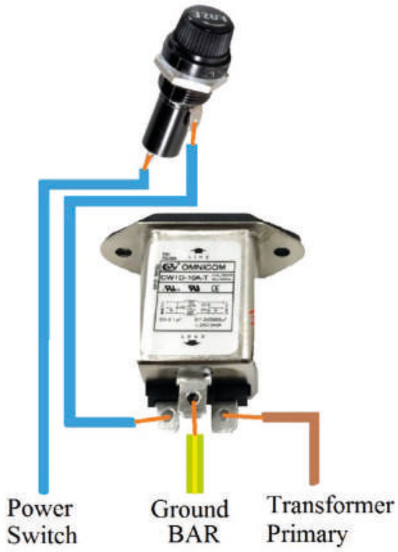
The inside partial chassis layout and wiring (BOTTOM VIEW)



Some details:

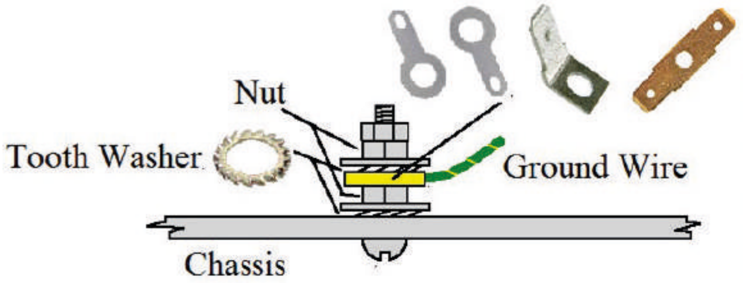


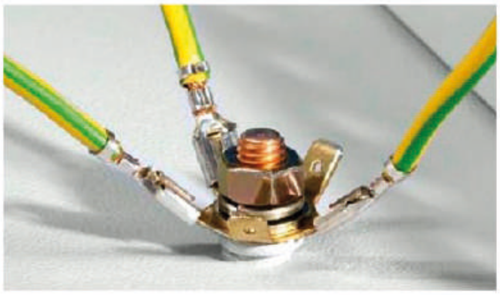
1. *The cinch connector of the input signal is wired to the volume potentiometer using an audio shielded cable. The shield is connected to the ground bar on the side of the audio cable that is closer to the potentiometer. The potentiometer ground pin is also connected to the ground bar. The middle pin of the potentiometer is connected to the control grid pin of the input tube socket.*



2. *Mains circuit: Primary coil of the transformer is connected to the mains connector or EMI filter. One end of the primary coil is directly wired to one EMI output pin. The other end of the primary coil is wired to the other EMI output pin via a power switch and mains fuse connected in series (EMI output pin is connected to one fuse holder pin. The other fuse holder pin is connected to the power switch pin. The other power switch pin is connected to the other end of the primary coil).*

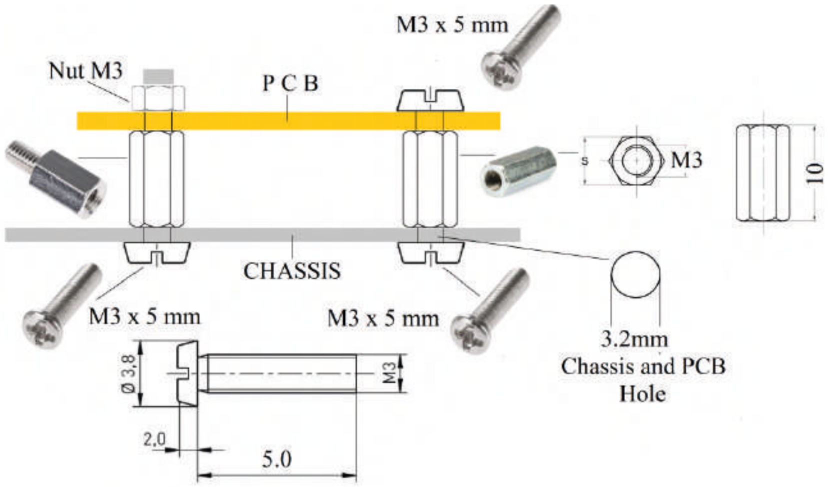
3. *For safety reasons, it is very important to implement the chassis grounding very carefully.*
   In order to make good mechanical and electrical contact between the metal chassis and the ground wire via the contact screw, it is necessary to use a special toothed washer. The area of the chassis where the tooth washer is placed must be cleaned of any paint or metal surface protection (or any material of low electrical conductivity) in order to make very good electrical contact between the tooth washer and the metal chassis.



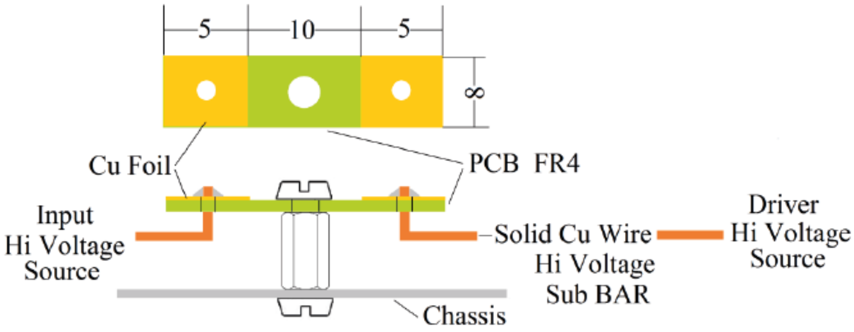


4. *The chassis ground point is connected to the ground bar via the RC circuit:*
   parallel connected resistor 10 Ω /5W and 0.22µF / 630V MKS capacitor.

8. *PCB mounting to the chassis (suggestion)*



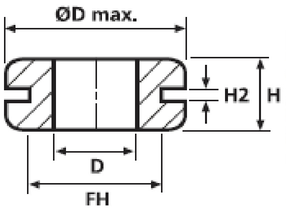
9. *Suggestion for making a sub-bar (auxiliary bar) using a piece of PCB and uninsulated solid Cu wire in order to carry and make electrical contact of electronic components connected between the socket pins of the tube and other parts or circuits of the amplifier such as high voltage power supply for example (electronic components of the input and driver stage anode circuit such as resistors can be connected to the anode pin of the tube socket and sub-bar).*

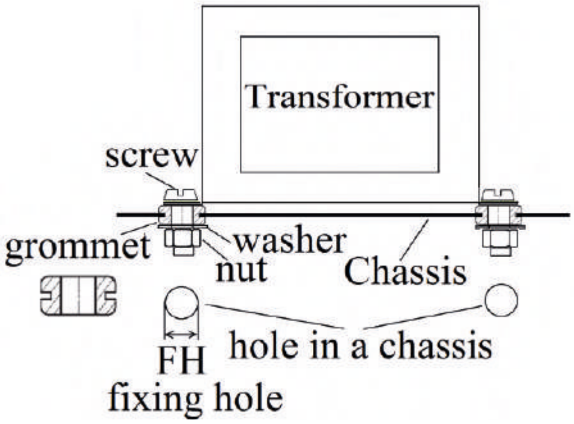


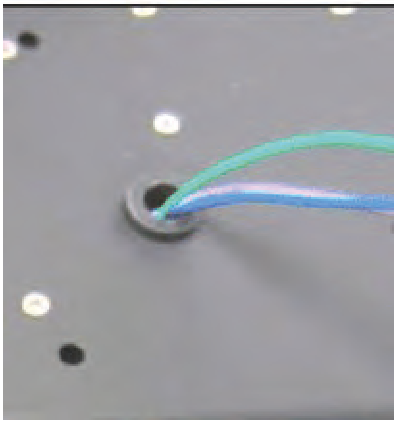
A few more suggestions

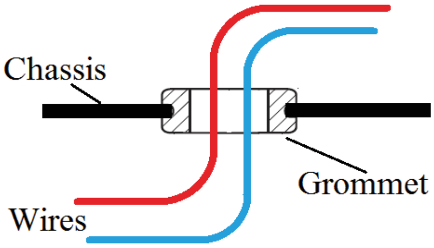
- *Mounting the transformer to the chassis*
  One way to minimize the transfer of transformer vibrations to the chassis is to mount the transformer to the chassis via some type of vibration absorber such as rubber washers, rings and cable grommets.

Ø Dmax [mm]	***Fixing Hole Ø FH***	Ø ***D***	***Height H***	***Height H2***
9.5	*6.4*	*4*	*5.6*	1.6
14.5	8	5	6.5	1.5
14	10	6	6.4	1.5
14	9.5	8	8	1.5
17.5	12	10	6.8	2





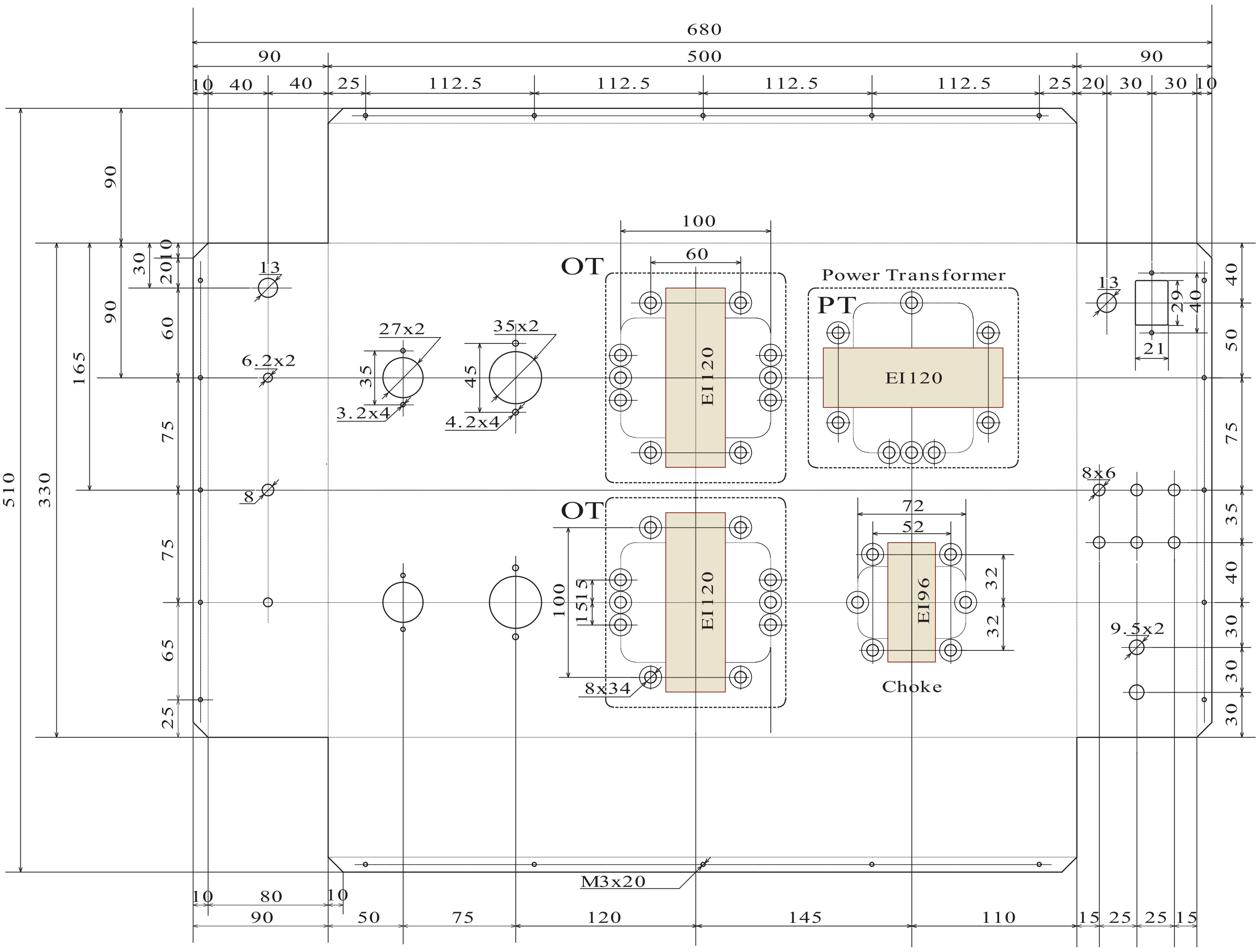




Placing cables and wires through chassis holes
To protect electrical cables and wires that pass through the holes in a chassis from a possible mechanical damage of the wire insulation and short circuit to the chassis, a rubber grommets can be inserted into the holes. Cables or wires pass through the grommets.

 # It is especially recommended for use in wiring hi voltage power supply and wiring of anode circuits of the amplifier stages, as well as for placing wires for wiring the primary output transformer circuit.

Example: Technical drawing for chassis sheet metal machining.



Material:

\# 1.2 mm cold or hot rolled steel (most common) or stainless steel (low magnetic, *chrome alloy steel*).
Cut and drill holes → Bend sheet metal and weld edges at all corners → Polish all welded corners.
(If the cold or hot rolled steel is used, the chasses must be protected from atmospheric affects by an electrochemical process or painted it with some thermostable paint).

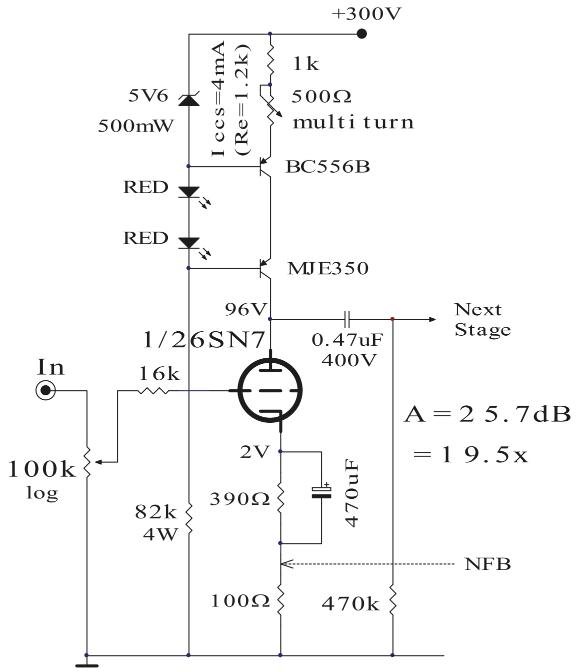
The above example of amplifier design should be understood only as a set of basic technical and practical procedures and operations in the process of designing SE amplifiers that can help the designer in the process of designing amplifiers.

Every designer, depending on theoretical and technical knowledge, skills, ideas and imagination, can, through theoretical elaboration, calculation and experimentation, find a constructive solution that satisfy his standards and perception of a good audio amplifier and the sound it produces.

There are no limits to the designer's imagination and desire for new constructions and the fidelity of the reproduced sound.
Some suggestions and examples from practice and ideas for upgrading the above design example:

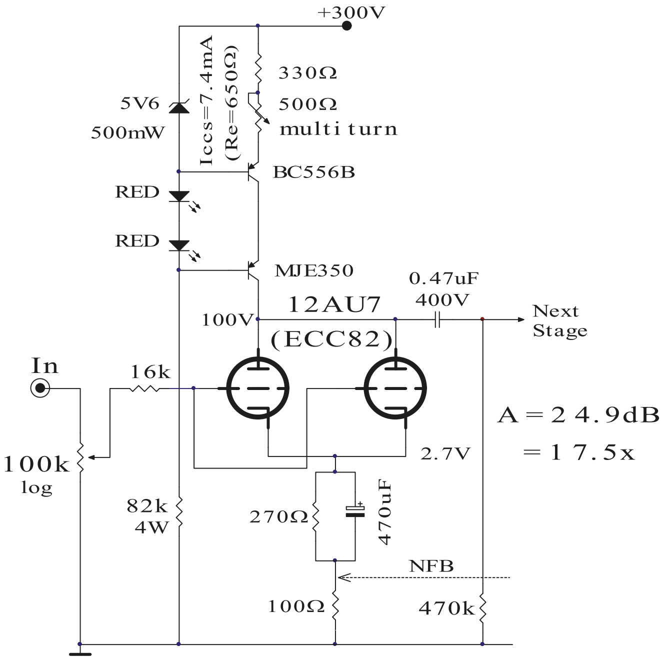
Input stage circuit:

Example1.



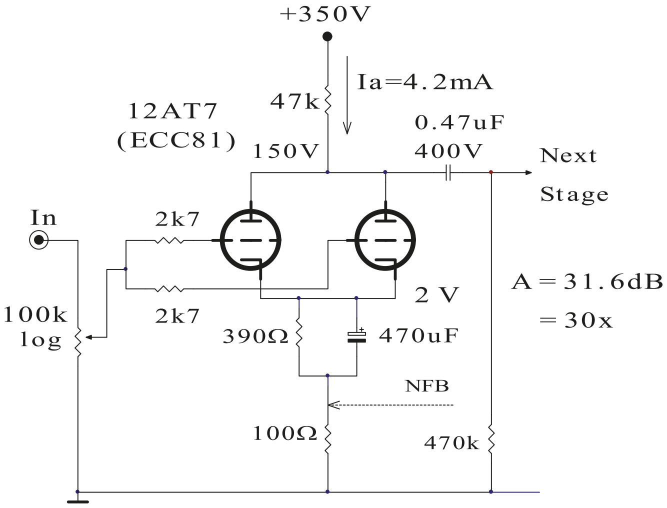
* Noval version: 6CG7 is equivalent to 6SN7
Also, 12AU7, slightly less amplification:
A = 24.9 dB (17.5 x)
$I_{CCS} = I_a = 3.7$ mA, Ua = 83 V, Uk = 1.8 V

Example 2:

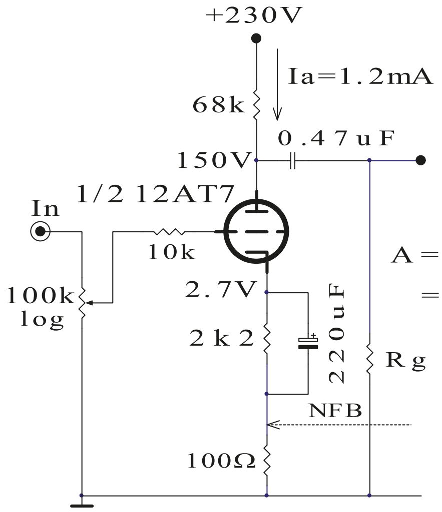


* Amplification is the same as with one tube
Output impedance is half that of one tube
Octal version: 6SN7, A = 25.7 dB (19.5 x)

Example 3:



Example 4:



Example 5:
Low noise small signal pentode EF86

Capacitances
Capacités
Kapazitäten

C_{g1} = 3,8 pF
C_a = 5,3 pF
C_{ag1} < 0,05 pF
C_{g1f} < 0,0025 pF

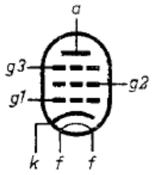
Heating : indirect by A.C. or D.C. series or parallel supply
Chauffage: indirect par C.A. ou C.C. alimentation série ou parallèle
Heizung : indirekt durch Wechsel- oder Gleichstrom; Serien- oder Parallelspeisung

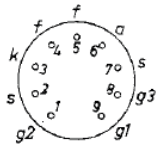
V_f = 6,3 V
I_f = 200 mA

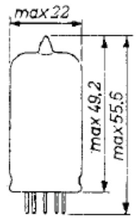
Typical characteristics
Caractéristiques types
Kenndaten

V_a = 250 V
V_{g3} = 0 V
V_{g2} = 140 V
V_{g1} = -2 V
I_a = 3,0 mA
I_{g2} = 0,6 mA
S = 2 mA/V
μ_{g2g1} = 38
R_i = 2,5 MΩ
R_{eq} < 0,1 MΩ

Dimensions in mm
Dimensions en mm
Abmessungen in mm



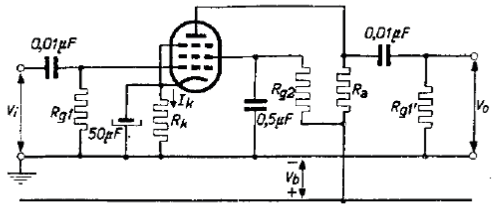




Base, culot, Sockel: NOVAL

Manufacturer's (Philips) recommendation:

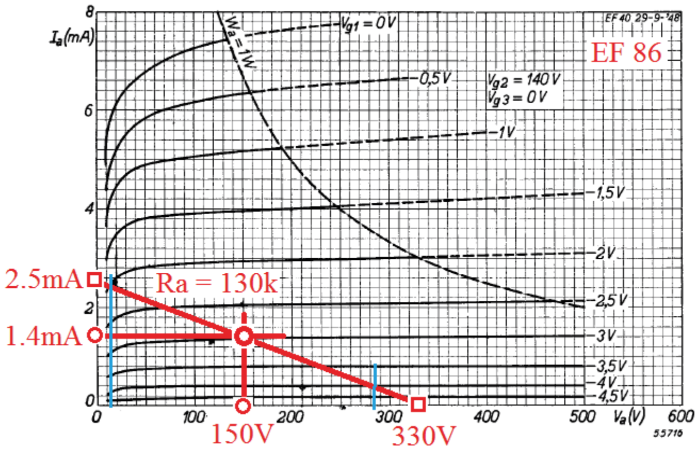
Operating characteristics as A.F. amplifier
Caractéristiques d'utilisation en amplificatrice B.F.
Betriebsdaten als NF-Verstärker

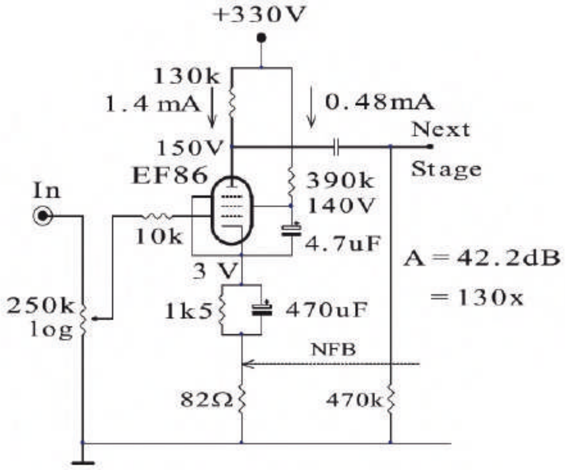


R_a = 100 kΩ; R_{g1}' = 330 kΩ; d_{tot} = 5%

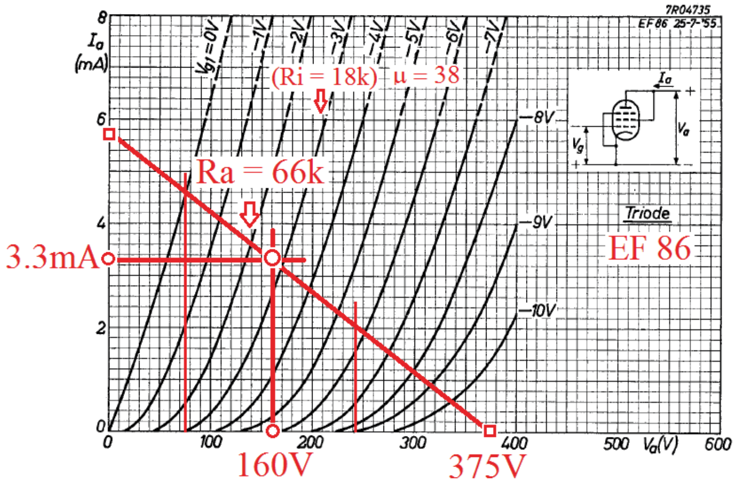
V_b (V)	I_k (mA)	R_{g2} (MΩ)	R_k (kΩ)	V_o/V_i 1)	V_o (Veff)
400	3,3	0,39	1,0	124	87
350	2,9	0,39	1,0	120	75
300	2,5	0,39	1,0	116	64
250	2,1	0,39	1,0	112	50
200	1,7	0,39	1,0	106	40
100	1,0	0,47	1,5	95	22

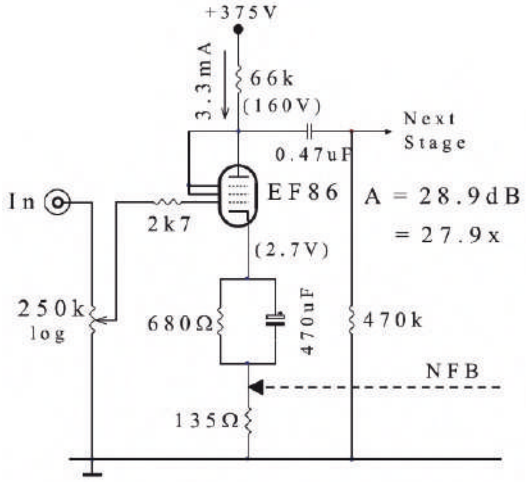
Input stage circuit:
High amplification input stage circuit - small signal pentode.





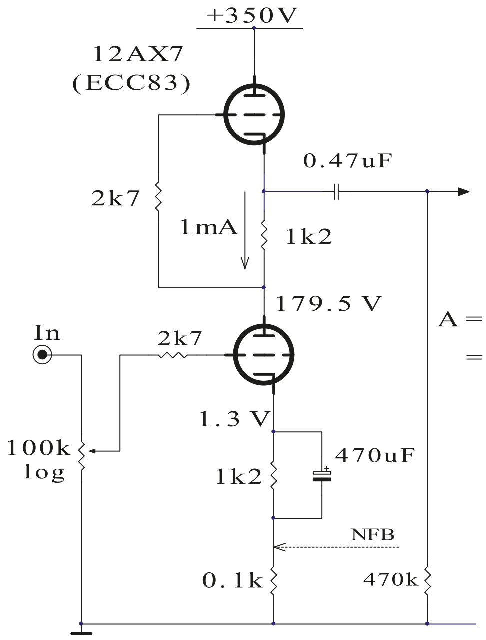
Example 6:
Low distortion input stage. EF86 triode mode.



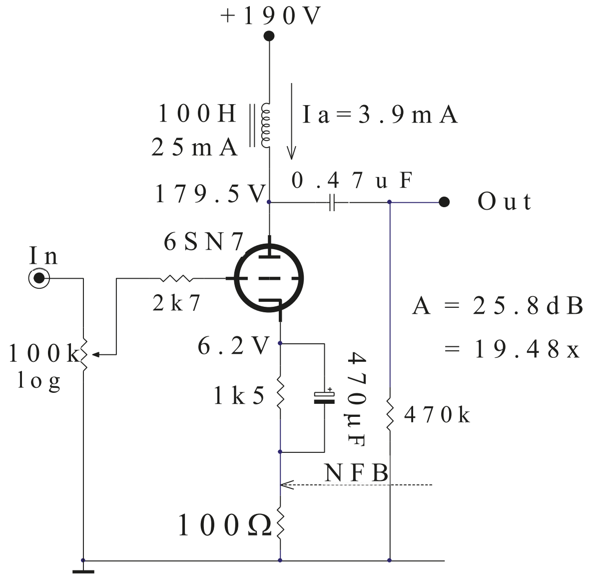


 # Same circuit suitable for DC coupling with the next stage (low anode voltage):
U_{ab} = 130 V, R_a = 100k, R_k = 1k8 II 100 μF + 100 Ω, U_{g-k} = - 1.4 V, I_a = 0.7 mA,
U_a = 60 V, A ≈ 25.

Example 7:
SRPP configuration

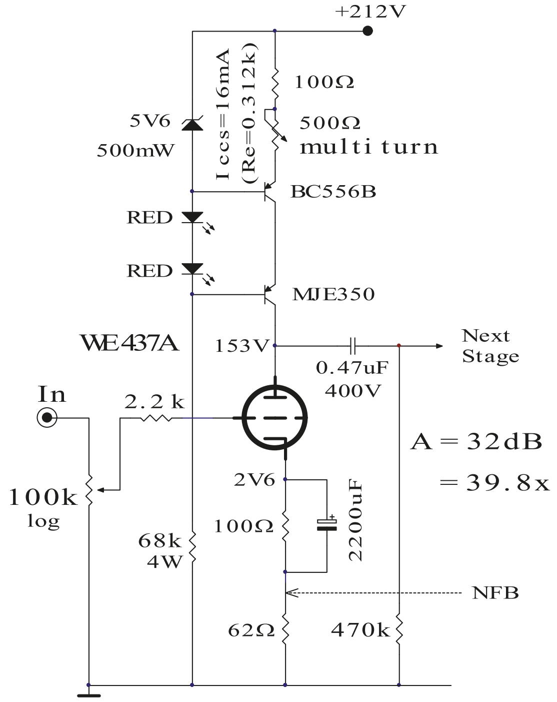


Example 8:
Inductive (choke) anode load



 # Low amplification: 6SN7, R_k = 1k5, I_a =3.8 mA, $U_{g\text{-}k}$ = –6 V. ***A = 23dB (15×).***

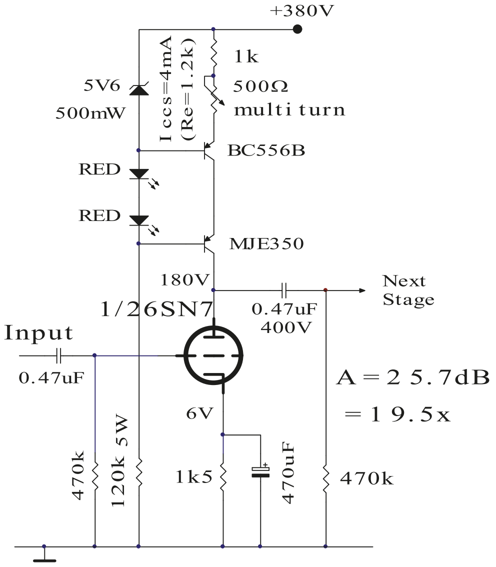
Example 9:
Author's input stage circuit used in SSE 2 × WE 300B



Driver stage

Driver stage design tips:

The driver stage should provide a signal to drive the amplifier output stage. In practical applications, the amplitude of the driver signal can be very high – up to several tens or even hundreds of volts peak to peak. Therefore, it is necessary to pay attention to the choice of the load line and the quiescent point of the driver's stage tube. Their choice should be such as to provide sufficient space to the left and right of the quiescent point on the load line for the driver stage output signal. Therefore, when selecting the operating conditions of the driver tube, the quiescent point of the *driver tube is shifted to higher anode voltages to ensure handling of the driver output signal of high amplitude. As the gain of the driver tube is limited and not infinite, the input signal of the driver tube can also be high – from a few volts to a few tens of volts peak to peak. Therefore, the U_{g-k} of the tube must be high enough to handle the input signal of the driver stage. As the driver stage works as a voltage amplifier and not as a power amplifier, it is not necessary for the driver tube to be high power.*

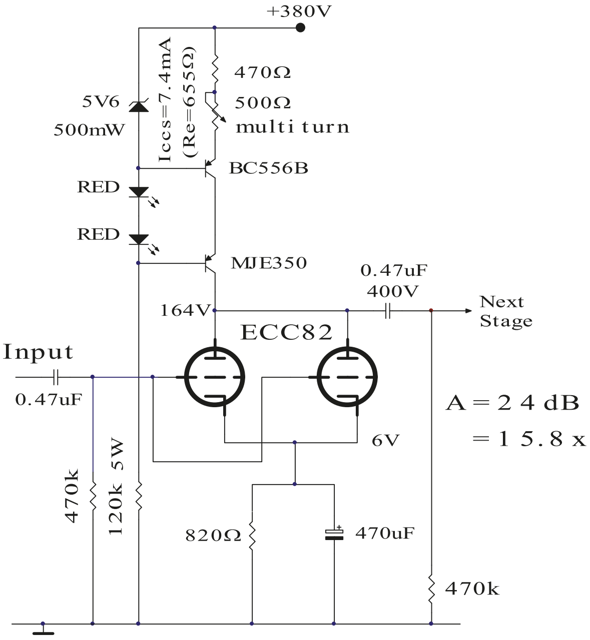


Example1:
Theoretically, the maximum amplification and the maximum amplitude of the output signal of the grounded cathode amplifier can be achieved if the anode load is infinite – load line is horizontal (I_a = constant). Therefore, in practice, CCS is sometimes used as an anode load.

Output voltage handling: over the 230 V_{P-P}

Noval: ECC82, I_{CCS} = 3.7 mA, R_K = 1k6;

A = 24 dB (15.8×)

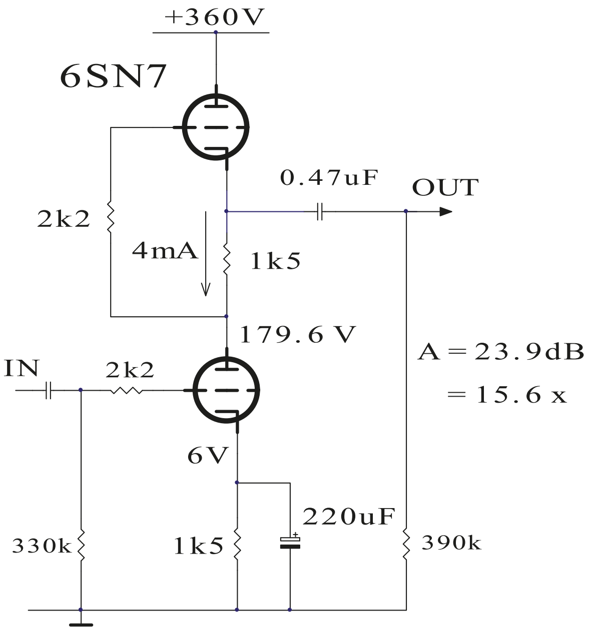


Example2:
Twice lower output impedance than in Example 1.

Octal: 6SN7, I_{CCS} = 8mA, R_K = 750 Ω

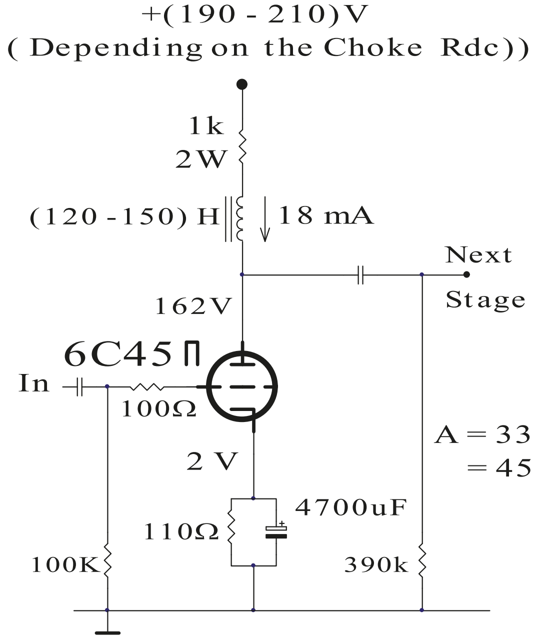
A = 26 dB (19.9 x)

Example 3:



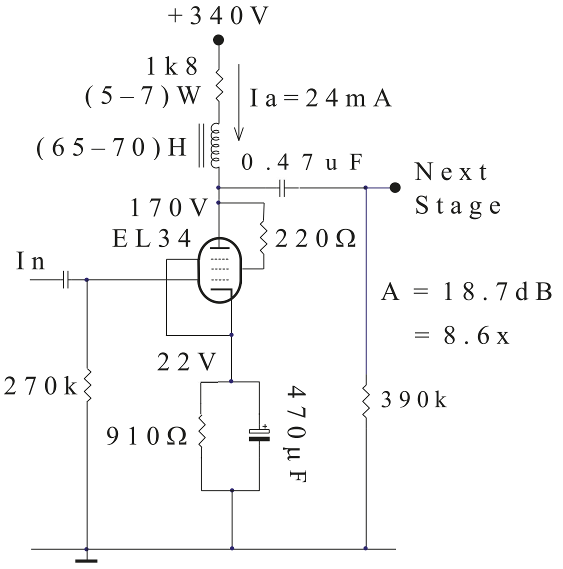
Note:
Medium amplification
Low output impedance

Example 4:



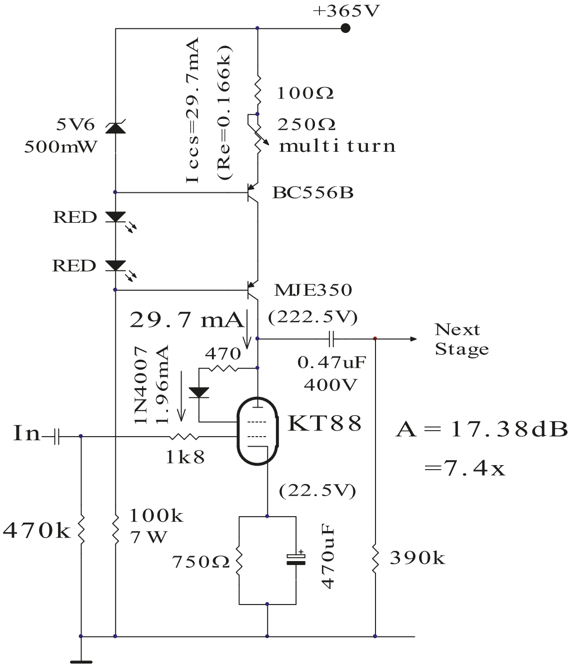
Note:
High amplification
Low output impedance

Example 5:
Triode mode power pentode
inductive (choke) anode load



Note:
Low amplification
Low output impedance

Example 6:
Triode mode power beam tetrode
CCS anode load



High End Push-Pull Vacuum Tube Power Amplifier for home music listening.

Design flow chart:
Phase 1. Designer's idea - decision:

Design requirements:

Output Power:	P_{out} = 2 × 30 W
Output Impedance:	R_L = 4 Ω
Input Sensitivity:	U_{in} = (500 - 780) mV
Input Impedance:	R_{in} = 250 kΩ
Amplitude Characteristic:	B_{min} = (20 Hz – 20 kHz)

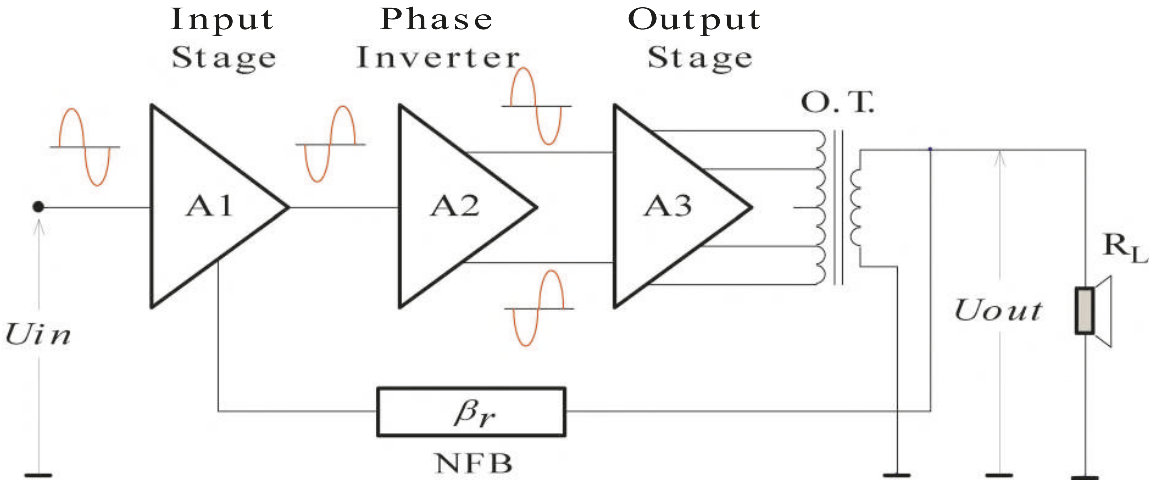
1. Output Stage
***# Ultra – linear push pull output stage** delivers much more power than an output stage that uses tetrodes (pentodes) connected as triodes and has a lower harmonic distortion (and outputimpedance) than the output stage that uses tetrodes (pentodes).*

2. Phase splitter
***# Cathode coupled phase splitter**.*
It can deliver high level (amplitude) output signal.
Grid circuit at elevated potential (relative to ground) allows the application of DC coupling tothe previous (input) stage (no coupling capacitor).
Low harmonic distortion and acceptable symmetry of output signals and output resistances.

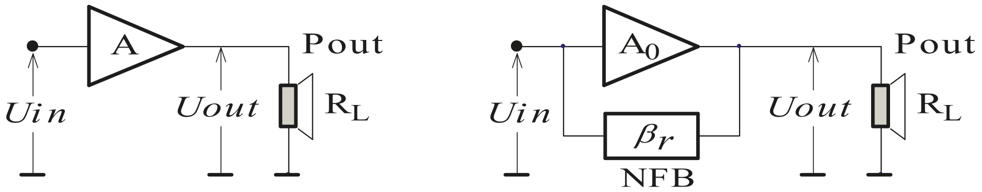
3. Input stage
***# Grounded cathode**.*
High signal amplification and high input resistance. Easy application (injection) of NFB.
Low nonlinear (harmonic) distortion (THD).

4.Global NFB.
***# Amplifier with negative feed – back** has lower nonlinear (harmonic) distortion (THD).*
Increased the bandwidth of the amplifier. Amplitude characteristic is almost frequency independent.

Phase 2. Block diagram:



Voltage amplification:



Input data (design requirement): $P_{out} = 30\ W$ *and* $R_L = 4\ \Omega$*:*

$$P_{OUT} = \frac{U_{OUT}^2}{R_L} \quad \rightarrow \quad \boldsymbol{U_{OUT}} = \sqrt{P_{OUT} \times R_L} = \sqrt{30W \times 4\Omega} = \sqrt{120} = \mathbf{10.95V}$$

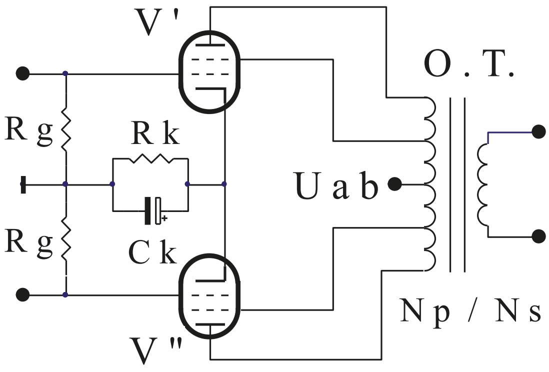
Total amplification:

$$\boldsymbol{A} = \frac{U_{OUT}}{U_{in}} = \frac{10.95V}{(0.5 \div 0.78)V} = (\mathbf{14} \div \mathbf{21.9})x \quad or, (\mathbf{22.9} \div \mathbf{26.8})\ dB$$

Voltage amplification without NFB:
Taking into account harmonic distortion, amplitude characteristic and stability of the amplifier, an ***NFB*** *of* ***20 dB*** *or ×* ***10*** *is applied, so the voltage amplification of the amplifier without NFB can be calculated:*

$$\frac{A_0}{A} = 10 \quad \rightarrow \quad \boldsymbol{A_0} = 10 \times A = 10 \times (14 \div 21.9) = (\mathbf{140} \div \mathbf{219}\ \boldsymbol{x})\ or, (\mathbf{42.9} \div \mathbf{46.8})\ dB$$

Step 3. Ultra Linear Push Pull Output Stage:



- Choice of output tubes:

KT 88 *is a well- known beam power tetrode still in production with many tube manufacturers.*

- Choice of bias type:
The main advantage of automatic bias over fixed bias is that the negative consequences caused by different characteristics of the tubes are minimized by the self-balancing effect.

- Operating conditions:
Based on a detailed analysis of the characteristics of KT88 as well as based on the analysis of the load line, some basic initial parameters of the output stage calculation can be defined:

Quiescent point:

- $U_{a\text{-}k} = 380\ V$
- $I_a = 75\ mA$
- $U_{g\text{-}k} = -40\ V$

Class: AB

- Ultra linear 43%
- $R_{a\text{-}a} = 6600\ \Omega$

KT88, internal (plate) resistance and gain under the above operating conditions:

- R_i = 2k4 (approximately)
- μ (total) ≥ 15 (approximately)

Calculation of expected output power:

$$\boldsymbol{P_{OUT\,(\max)}} = \frac{2 \times \left(\mu \times U_{g\,(\max)}\right)^2}{(2 \times R_i + R_{a-a})^2} \times R_{a-a} = \frac{2 \times (15 \times 40)^2}{(2 \times 2400 + 6600)^2} \times 6600 = \mathbf{36.5\ W}$$

$U_{g\,(max)}$ – maximum amplitude of drive signal ($\leq U_{g\text{-}k}$)

Voltage amplification:

$$\boldsymbol{A} = 2 \times \left(\mu \times \frac{\frac{R_{a-a}}{2}}{R_i + \frac{R_{a-a}}{2}} \right) = 2 \times \left(15 \times \frac{\frac{6600}{2}}{2400 + \frac{6600}{2}} \right) = \mathbf{17.36}\ \boldsymbol{x}; or, (\mathbf{24.79\ dB})$$

The total voltage amplification of the UL Push Pull Output Stage including the Output Transformer is:

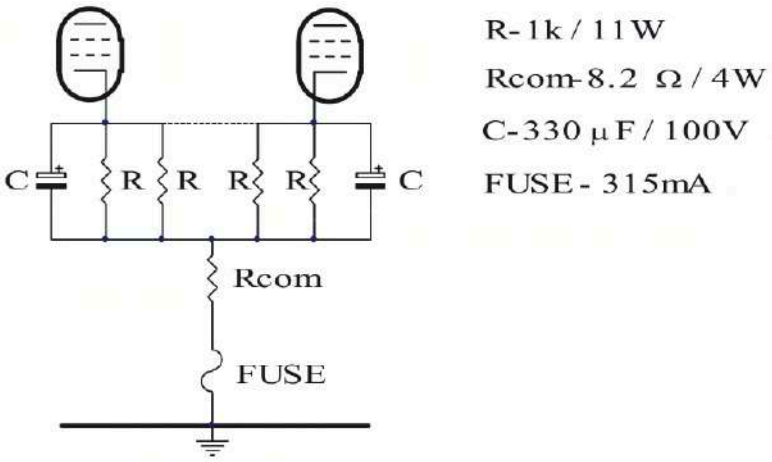
$$\boldsymbol{A_3} = A \times \sqrt{\frac{R_L}{R_{a-a}}} = 17.36 \times \sqrt{\frac{4}{6600}} = \mathbf{0.427}\ or - \mathbf{7.39\ dB}$$

Calculation of automatic bias resistor (common cathode resistor – one common cathode resistor of connected cathodes of both tubes):

$$\boldsymbol{R_k} = \frac{U_{g-k}}{2 \times I_a} = \frac{40}{2 \times 0.075} = \mathbf{266\ \Omega}$$

$$\boldsymbol{P_{R_k}} = R_k \times (2 \times I_a)^2 = 266 \times (2 \times 0.075)^2 = \mathbf{5.985\ W}$$

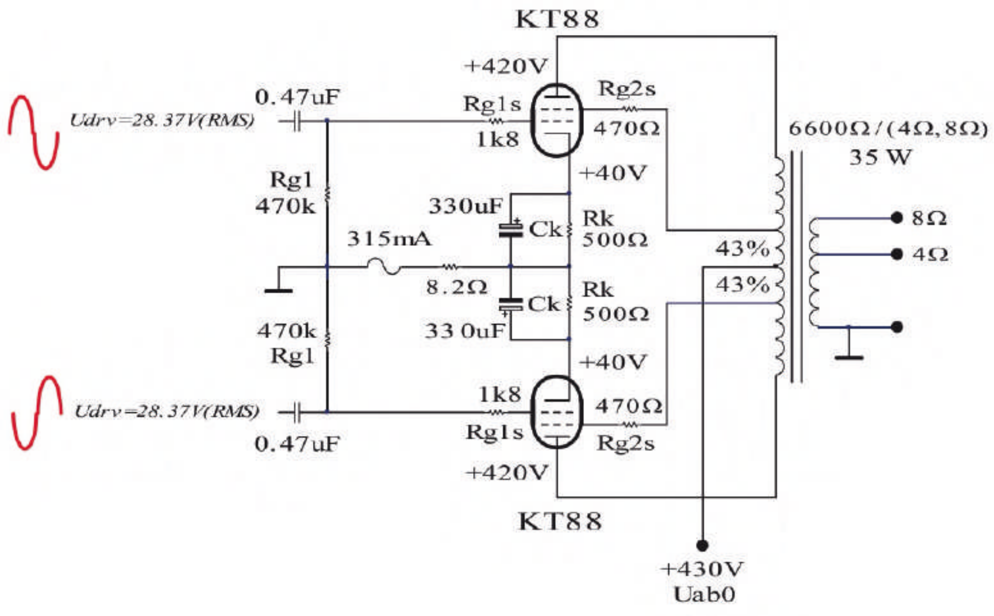
*In practice, R_k can be made using four 1 kΩ/(7 W÷11 W) resistors connected in parallel (4 || 1 kΩ / (7 W÷11 W) = 250 Ω / (28 W÷44 W) and one **8.2 Ω / 4 W** resistor connected in series.*



 # Note: Some KT 88 manufacturer recommend the use of separate cathode resistors.
Following this recommendation, the cathode resistor of each tube can be made using two resistors connected in parallel: (2 × 1 kΩ / (7 W÷11 W – 500 Ω)). The cathode resistors of the tubes are grounded via a common 8.2 Ω / 4 W resistor and a fuse that acts as a simple overcurrent protection.

Reminder:
The effect of NFB (caused by cathode impedance which reduces the gain and increases the output impedance of the common cathode stage) can be partially or fully eliminated by bypassing the cathode bias resistor $\boldsymbol{R_k}$ to the ground for AC signals – parallel connection R_K and capacitor C_K.

By adding grid resistors (R_g = 470 kΩ, recommended for KT88 by the manufacturers) and grid stopper resistors, the electrical diagram of the output stage is completed:



$R_{L\ a\text{-}a} = 6600\Omega$

$UL = 43\%$

$P_{OUT} = 36.5W$

$A_3 = 0.427$

$U_{drv} = 28.37V_{RMS}$

Problem of mismatched output tubes

The basic principle of operation of the PP output stage is based on the principle of symmetry or, in other words, identical characteristics of the output tubes. But in practice, it is impossible to produce two identical tubes. Tube manufacturers usually publish data on deviations of the characteristics of the produced tubes from the characteristics of the tubes published in the data sheets (tolerances). Tolerances of ±5% and ±10% are common for series-produced tubes. So-called pairs (or quartets) of matched tubes with smaller tolerances can be found on the market. In practice, a tolerance of ±1.5% in tube pair characteristics is quite acceptable for use in Hi End PP amplifiers.

Consequences of the difference in the characteristics of the tubes (mismatched tubes) on the operation of the PP output stage:

- *Current unbalance in a PP output stage, the quiescent point currents of the output tubes are different. At the same time, this means that the currents flowing through the halves of the primary windings of the output transformer are unbalanced. Unbalance of DC currents flowing through the output transformer can cause saturation of the iron core and lowering the inductance of the Primary of the transformer.*

  Consequences of a low inductance of the Primary of the output transformer to the amplifier characteristic:
    - *Poor amplitude characteristic (frequency band) of the amplifier at low frequencies (higher cut-off low frequency)*
    - *Higher harmonic distortion at low frequencies (especially at high power levels).*

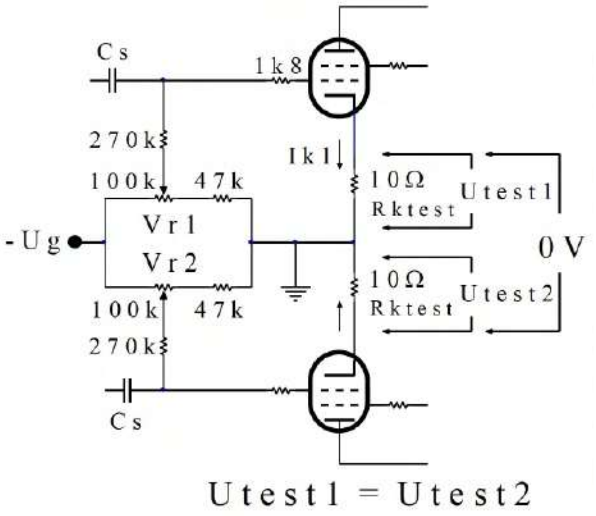
- *Increased harmonic distortion of even–order harmonic in general.*

Minimizing the problem of mismatched output tubes

A carefully designed bias circuit can minimize the consequences of mismatched tubes in practice.

- Fixed bias of output tube

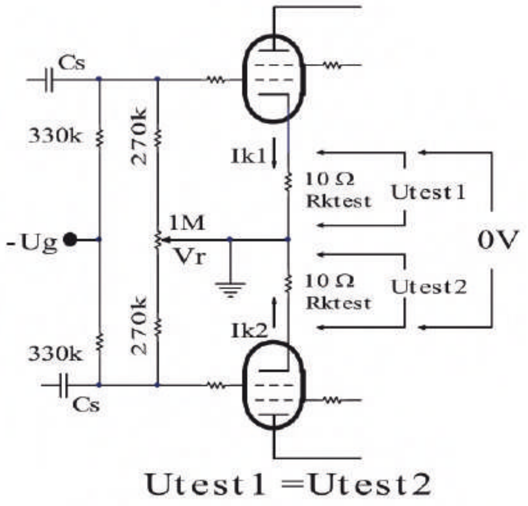
  a) Each tube has its own fixed bias circuit.



By varying Vr1 and Vr2, the grid bias is adjusted so that the current flowing through the one tube is equal to the current flowing through the other tube. The adjustment process is usually iterative — consequently, the process of adjusting the balance of the currents (adjusting with Vr1 and Vr2) should be repeated several times until the maximum possible balance of the currents is reached:

Ik1 = Ik2; Utest1 = Utest2;

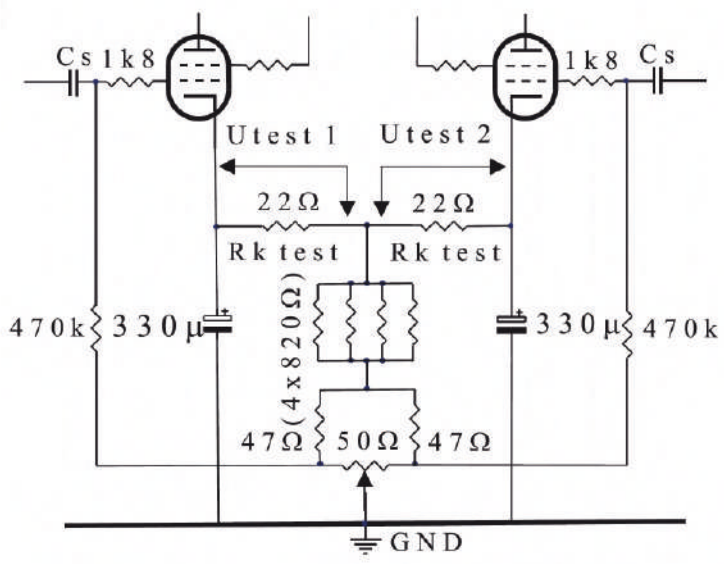
b) Bias balancing



 # Note (for both circuits):

Rk test1 = Rk test2 , low-tolerance 1% resistor.

- Automatic bias of output tubes
  A very effective balancing circuit applied in a Williamson amplifier (modified).



Another component of the PP output stage plays a significant role in fulfilling the conditions of symmetry of the PP amplifier — the output transformer.

Output Transformer

As it is necessary for the output tubes to be paired, it is also important that both halves of the Primary windings have the same characteristics: DC and AC. In other words, both halves of the Primary should have the same DC resistance and the same inductance. The condition of symmetry of the halves of the Primary windings is relatively easily to realize in practice by careful construction of the output transformer — by winding a symmetrical pair of coils.

 # Note on the choice of the transformer core material.

- PP output transformer using E / I laminate core can tolerate small DC currents imbalance. No matter how fine packed the E and I laminates are, there is always an air gap between them and no matter how small it affects the magnetic properties of the transformer core.
- PP output transformer that uses a toroidal core is much more sensitive to DC currents imbalance – almost absolute balance of DC currents is required.

 # General notes

- *In the case of PP amplifier designed above (automatic bias), the imbalance of DC currents of the output tubes of several mA is acceptable (1 - 3) mA and no additional balancing circuit is required.*
- *If the imbalance of DC currents of the output tubes of the PP amplifier is unacceptable (unpaired output tubes) and it is necessary to use a DC currents balancing circuit of the output stage tubes, in the process of adjusting the complete amplifier one of the first adjustment should be balancing of DC currents of the output stage.*

- *After the procedure of balancing the DC currents of the output tubes is completed, the procedure of balancing the AC signals by balancing the output signals of the phase splitter stage (if it is possible) follows. DC balancing of the output stage does not mean that the characteristics of the output tubes (gain, internal resistance) are paired, on the contrary, they always differ, which results in unequal amplification of phase splitter output signals and imbalance of the AC signals amplitudes of the primary coils of the output transformer.*

- *In fact, the best method for AC balancing of PP amplifiers is to adjust the amplitudes of the output signals of the phase splitter stage to obtain minimal harmonic distortion at the output of the amplifier. If there is no technical possibility to adjust the levels of the output signals of the phase splitter, the reduction of the harmonic distortion of the amplifier can be tried to by replacing the locations of the output tubes.*

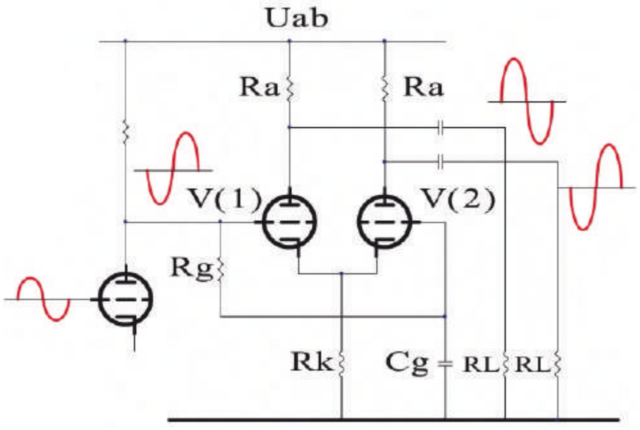
- *However, the best technical characteristics of PP amplifiers are achieved by screening and selecting of output tubes - maximum pairing of output tubes according to all parameters — DC and AC (gain, transconductance, internal resistance) as well as using a quality output transformer with perfectly symmetrical halves of the primary windings.*

Phase-splitter

Phase-splitter circuit:

- *Cathodyne phase inverter (concertina phase inverter).*
- *Long Tail Pair (cathode coupled) inverter.*
- *Schmidt inverter.*

In this example, the choice is a ***Long Tail Pair****.*

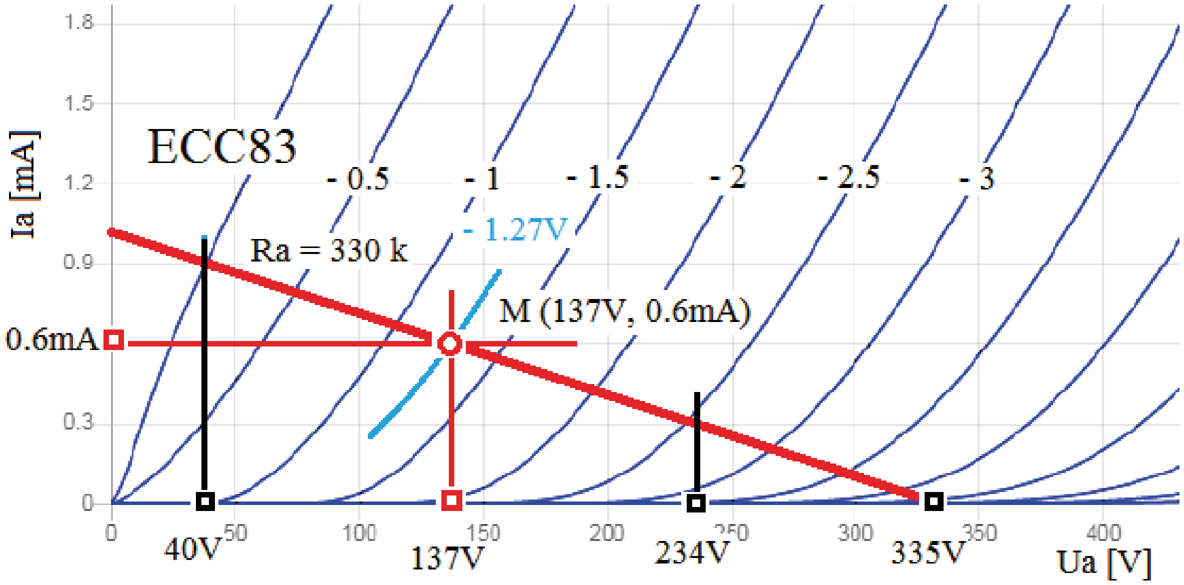


- *The input and output signals of the first tube are opposite in phase.*
- *The input signal of the first tube and the output signal of the second tube are in phase.*
- *The amplification of the first and second tubes is not equal (this difference is* $\frac{A_{11}}{A_{12}}=1+\frac{R_i+R_a}{R_k\times(\mu+1)}$ *and is smaller if the* $(\mu+1)xR_k \gg (R_i+R_a)$**). In order to minimize this difference, it is necessary to use tubes with high* μ *.*

(In order to make the amplifications of both outputs equal, it is necessary to lower the anode resistor of the inverted output: $R_{a1}=\frac{R_a}{1+\frac{R_i+R_a}{R_k\times(\mu+1)}}$*).*

The ***ECC83*** *is the double triode with the highest* μ *(100) currently in production.*
As can be seen from many examples and applications as well as from graphical analysis of the load line, the operating conditions of ECC83 can be:

- $U_{a-k} = (135 \div 140)\ V$
- $I_a = (0.6 \div 0.65)\ mA$
- $U_{g-k} = -(1.2 \div 1.3)\ V$
- $R_a = 330\ k\Omega$



As can be seen from the graphical analysis, the load line and quiescent point selecting in this way provide high level output signals (almost ± 97V), sufficient to drive the output tubes.

The common cathode is at an elevated potential (and also the grids of both tubes) relative to the ground, so it is possible to make a direct coupling between the input stage and the cathode coupled phase-splitter. In this example, taking into account the maximum allowable voltage between the cathodes and the tube heaters, as well as the possibility of making a direct coupling to the input stage, the choice is that this potential is 68 V to 70 V.

Calculation of common cathode resistor:

$$R_K = \frac{U_K}{I_{a1} + I_{a2}} = \frac{(68 \div 70)V}{2 \times (0.6 \div 0.65)mA} \approx \mathbf{56\ k\Omega}$$

If $\mathbf{R_a = 330\ k\Omega}$ *, then the amplification:*

Amplification of the first tube:

$$A_{11} = -\frac{\mu \times R_a'}{(R_i + R_a') \times \left[2 + \frac{R_i + R_a'}{R_k \times (\mu + 1)}\right]} \times \left[1 + \frac{R_i + R_a'}{R_k \times (\mu + 1)}\right]$$

$$= \frac{100 \times 193.8}{(62.5 + 193.8) \times \left[2 + \frac{62.5 + 193.8}{(100 + 1) \times 56}\right]} \times \left[1 + \frac{62.5 + 193.8}{56 \times (100 + 1)}\right] = \mathbf{38.6}$$

Where:

- R_i *= 62.5kΩ, internal resistance of ECC83.*
- $R_a' = \frac{R_a \times R_{g1\,out}}{R_a + R_{g1\,out}} = \frac{330k\Omega \times 470k\Omega}{330k\Omega + 470k\Omega} = 193.8\ k\Omega;$

  $R_{g1\ out}$ is the control grid resistor of the output tube (470 kΩ).

Amplification of the second tube:

$$A_{12} = -\frac{\mu \times R_a'}{(R_i + R_a') \times \left[2 + \frac{R_i + R_a'}{R_k \times (\mu + 1)}\right]} = \frac{100 \times 193.8}{(62.5 + 193.8) \times \left[2 + \frac{62.5 + 193.8}{(100 + 1) \times 56}\right]} = \mathbf{36.96}$$

The difference in the amplification of the first and second tubes can be minimized by using a 300kΩ resistor instead of the 330kΩ in the anode circuit of the first tube, or an amplification balancing circuit can be used.
In the following calculations, the amplification of the cathode coupled phase splitter of $\boldsymbol{A_2 = 37 \times}$ *or* ***31.36 dB*** *is used.*

The high voltage *of the phase splitter supply has to be:*

- common cathode voltage + voltage across the tube + voltage drop across the anode resistor:

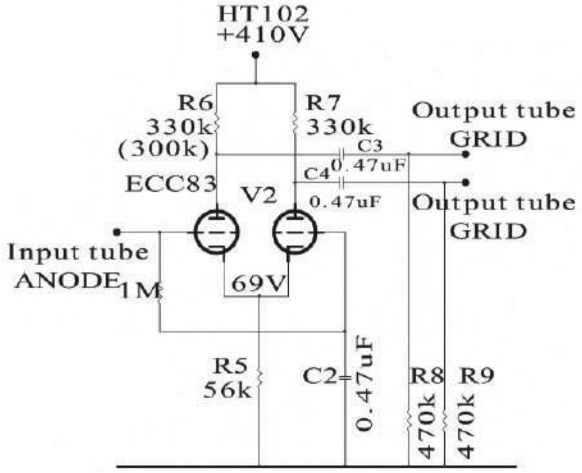
$U_{ab2} = 68 + 135 + 0.625x330 \approx \mathbf{410\ V}$

Coupling capacitors *between phase-splitter and output stage:*

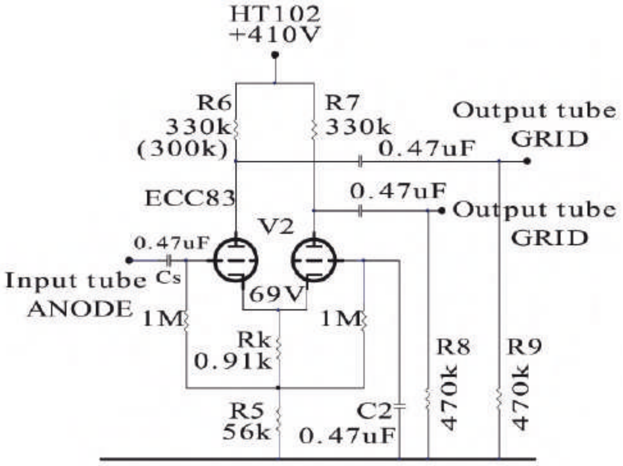
Using the equation $f_{min} = \frac{1}{2\times\pi\times C_S\times\left(R_g+\frac{R_i\times R_a}{R_i+R_a}\right)}$ *and the required* $f_{min} \le 1\ Hz \rightarrow C_S = 0.47\ \mu F$

The total amplification of the driver and output stage is:

$$A_{drv+out} = A_{drv}\times A_{out} = A_2\times A_3 = 37\times 0.427 = 15.8x\vee, 23.97\,dB$$

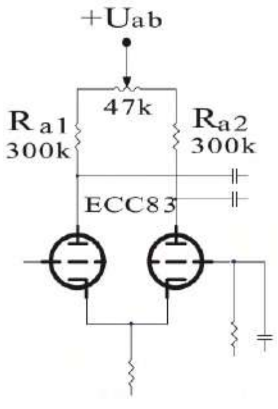


DC COUPLING

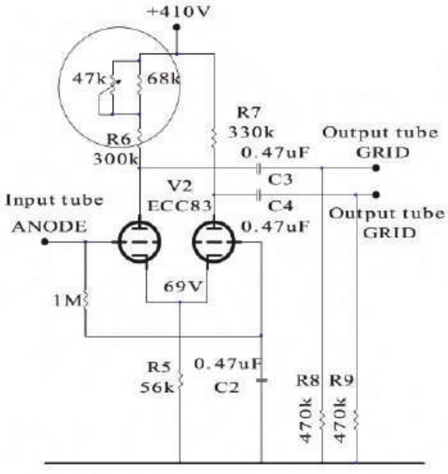


RC COUPLING

The characteristics of a long tail pair circuit is not only that the output signals are opposite in phase, but also that they differ in amplitude. Thus, the basic principle of the PP amplifiers configuration, symmetry, is violated. This amplitude inequality does not have to drastically affect the characteristics of the complete amplifier in each particular case. The principle of symmetry of PP amplifiers is also affected by the symmetry of the output stage (in some cases the relation between the asymmetry of the amplitudes of the output signals of the phase - splitter and the asymmetry of the output stage can be such that a high level of symmetry of the complete PP amplifier can be made). In practice, there are also no two identical electronic components (for example, no two vacuum tubes with identical static and dynamic characteristics). Therefore, complete symmetry of the output signals of the phase-splitter does not have to be imperative, on the contrary, the most common case in practice is that these signals differ in amplitude and that the amplifier has good characteristics.



The difference in amplification of the first and second tube can be minimized by connecting a 300 kΩ resistor to the anode circuit of a first tube instead of a 330 kΩ resistor (anode resistor of the other tube is 330 kΩ), or by using a 47 kΩ trim potentiometer to adjusting (balancing) output signals levels (terminals of the fixed ends of the trim potentiometer are connected to a 300kΩ anode resistor of each tube respectively and a wiper terminal is connected to the power supply U_{ab2}*, or, for example, using a circuit:*



SIGNAL BALANCING

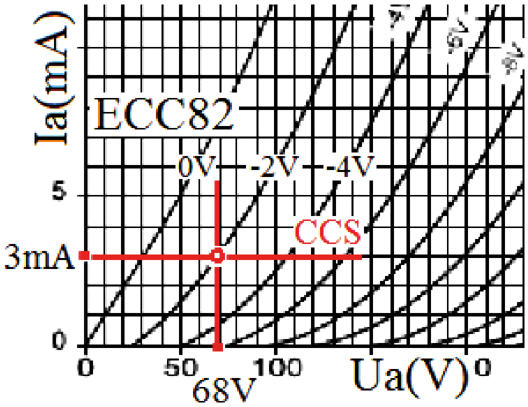
 # Note:
It is best to apply the AC balancing procedure on the complete amplifier, where it is not the most important that the signals of the amplifier stages are perfectly balanced, but to get a signal of minimal distortion (min THD) at the output of the amplifier.

Input Stage
In order to facilitate the process of selecting the configuration of the input circuit, as well as the process of selecting its electronic components, it is necessary to determine the minimum amplification of the input stage:

$$A_1 = \frac{A_0}{A_2 \times A_3} = \frac{(140 \div 219)}{37 \times 0.427} = (\mathbf{8.86 \div 13.86})\, x$$

or,

$$A_1[dB] = A_0 - (A_2 + A_3) = (42.9 \div 46.8) - (31.36 - 7.39) = (\mathbf{18.93 \div 22.83})\, dB$$



Another condition that must be met: the DC voltage at the output of the input stage must be close to the voltage at the cathodes (grids) of the phase splitter stage – around 68V (DC coupling condition).

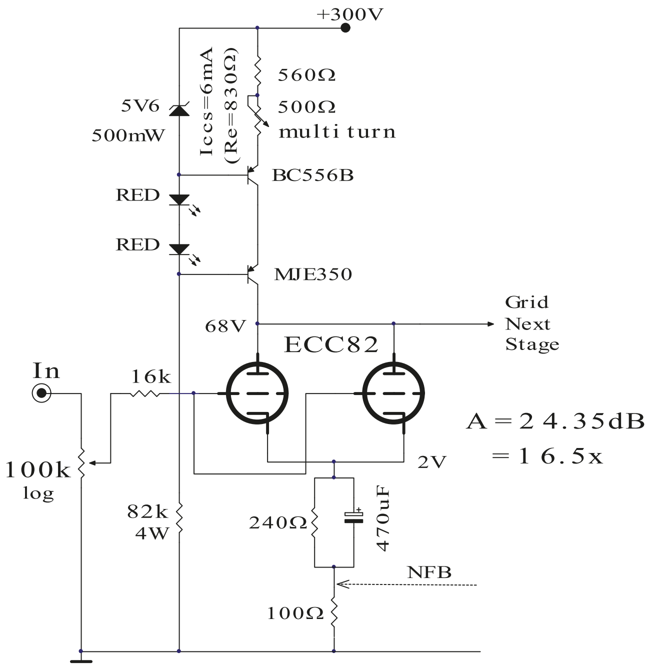
One example of an input stage that could meet these conditions is given as an example of an input stage of the SE amplifier – grounded cathode amplifier: a parallel connection of two halves of ECC82 with constant current source as anode load (the circuit should be adapted to the required conditions).

Two halves of ECC82 connected in parallel:

$I_a = 2 \times 3\ mA = 6\ mA$

A 6mA CCS can be used as the anode load.
A supply voltage of approximately +300 V can be used.
Cathode resistor: $R_k = 2\ V / 6\ mA \approx 340\ \Omega$

As the cathode circuit is part of NFB loop, the cathode resistor R_k *is designed as a series connection of a 100 Ω resistor and a 240 Ω resistor bypassed by a 470* µF capacitor.



The amplification of the input stage is almost equal to the μ of the tube (ECC82, μ = 17):

A_1 = 24.35dB (16.5 x).

The total amplification of the amplifier without NFB is:

$$\boldsymbol{A_0} = A_1 \times A_2 \times A_3 = 16.5 \times 37 \times 0.427 = \mathbf{260\ x}$$

$$\boldsymbol{A_0}[dB] = A_1 + A_2 + A_3 = 24.35 + 31.36 - 7.39 = 48.5dB$$

With applied NFB of 20 dB, the amplification is:

A = 26 ×** or **28.5 dB.

Expected input sensitivity:

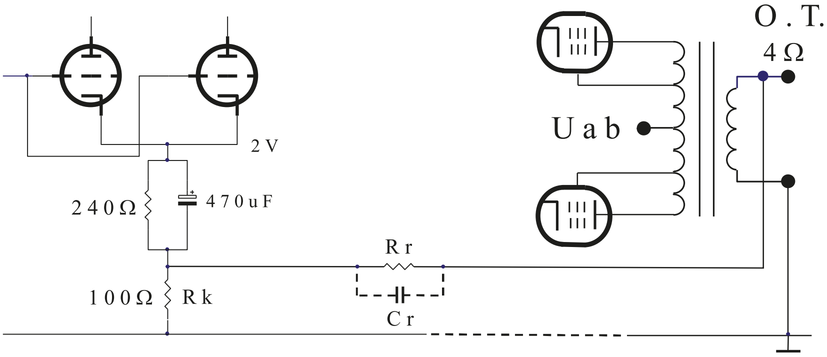
$$\boldsymbol{U_{in}} = \frac{U_{out}}{A} = \frac{10.95V}{26} = \mathbf{421\ mV}$$

Negative Feed-back circuit

The level of NFB is determined based on compromise between the positive effects of NFB application on the characteristics that determine the quality of the amplifier (amplitude characteristic, THD, output impedance,…) and stable operation of the amplifier (compromise: performance – stability).

NFB of 20 dB (10 x) is acceptable:

$A / A_0 = 10$



*Via the voltage divider **R_k / (R_r + R_k)** portion of the output signal is applied to the input circuit (cathode circuit) and acts as a negative feed - back.*

The amplification of the amplifier with applied NFB is:

$$A = \frac{A_0}{1 + \beta \times A_0} = \frac{A_0}{1 + \frac{R_r}{R_k + R_r} \times A_0}$$

*Using the above equation, the resistor **R_r** can be calculated:*

$$\frac{A}{A_0} = \frac{1}{1 + \beta \times A_0} = \frac{1}{10} \quad \rightarrow \quad \boldsymbol{R_r} = \frac{R_k \times (A_0 - 9)}{9} = \frac{100\Omega \times (260 - 9)}{9} = \mathbf{2789\ \Omega}$$

*Standard, **R_r = 2k7***

Capacitor C_r connected in parallel with resistor R_r makes the NFB circuit frequency dependent (mostly at a high frequencies) and it shapes the overall frequency response of the amplifier.

(The mathematical analysis of frequency-dependent NFB is quite complicated and goes beyond the scope of this book). In practice, the value of the capacitor (Cr) is determined experimentally by observing the shapes of the square wave signal at the output of the amplifier (more on this topic in another chapter).

The initial value of Cr can be determined using the equation:

$$C_r = \frac{1}{2 \times \pi \times f_r \times R_r}$$

f_r – cut-off high frequency (for quality amplifier: $f_r \geq 50$ kHz)

Example:
R_r =2k7 and f_r = 65 kHz:

$$C_r = \frac{1}{2 \times \pi \times f_r \times R_r} = \frac{1}{2 \times 3.14 \times 65000 \times 2700} = 907\,pF$$

 # Note:
If an 8Ω (instead of 4Ω) secondary of the output transformer is used, and if it is used as part of the NFB circuit, it is necessary to recalculate the amplification (A_0) of the amplifier without applied NFB and the value of the resistor R_r (by changing the impedances ratio of the output transformer, the amplification of the output stage (A_3) also changes).

$$\boldsymbol{A_3} = A \times \sqrt{\frac{R_L}{R_{a-a}}} = 17.36 \times \sqrt{\frac{8}{6600}} = \mathbf{0.604}\ or - \mathbf{4.37\ dB}$$

$$\boldsymbol{A_0} = A_1 \times A_2 \times A_3 = 16.5 \times 37 \times 0.604 = \mathbf{368.7\ x}$$

$$\boldsymbol{R_r} = \frac{R_k \times (A_0 - 9)}{9} = \frac{100\Omega \times (368.7 - 9)}{9} = \mathbf{3966\ \Omega}$$

*Standard, **R_r = 3k9***

$$C_r = \frac{1}{2 \times \pi \times f_r \times R_r} = \frac{1}{2 \times 3.14 \times 65000 \times 3900} = 628\,pF$$

Power supply
High voltage power supply

- Output stage:

  U_1 *= U_K + U_{a-k} + $U_{T\,dc}$ = 41V + 38V + $I_a \times R_{Prim.\,dc}$ = 41 V + 380 V + 0.075 mA × 103.5 = 428.7 V ≈* ***430 V***

  $R_{Primary\ dc}$ - DC resistance of the Primary winding of the output transformer (DC resistance of one half of the total primary winding).
  Push Pull Output Transformer, DC resistance of the primary in an example: "Push-Pull Output Transformer 35 W / 6600 Ω " is 103.5 Ω (DC resistance of one half of the primary winding).

  DC current of the output stage: ***I_{OUT}*** *= 2 × 75 mA =* ***150 mA (one channel)***

- Driver stage:

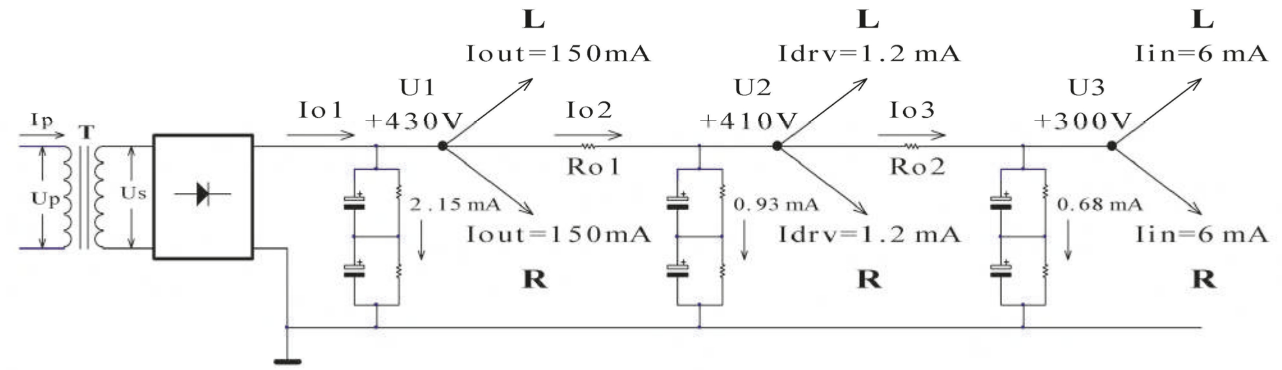
  U_{drv} = U_2 = 410 V
  I_{drv} = 1.2 mA (one channel)

- Input stage:

  U_{in} = U_3 = 300 V
  I_{in} = 6 mA (one channel)

Project request or designer's decision:
Common High Voltage Power Supply for both channel.

The concept of high voltage power supply:



According the concept schematic:

- I_{01} = 2.15 mA (current flowing through the balancing resistor of the smoothing capacitor) + 2 × I_{out} + 0.95 mA (current flowing through the balancing resistor of the smoothing capacitor) + 2 × I_{drv} + 0.68 mA (current flowing through the balancing resistor of the smoothing capacitor) + 2 × I_{in}

  $I_{01} = 2.15\ mA + 2 \times 150\ mA + 0.95\ mA + 2 \times 1.2\ mA + 0.68\ mA + 2 \times 6mA = \mathbf{318.8\ mA}$

Resistor calculation R_{01} and R_{02}:

- $I_{01} = 318.8\ mA$
- $I_{02} = 0.95\ mA + 2 \times 1.2\ mA + 0.68\ mA + 2 \times 6\ mA = 16.03\ mA$
- $U_{R01} = U_1 - U_2 = 430\ V - 410\ V = 20\ V$
- $R_{01} = \frac{U_{Ro1}}{I_{o2}} = \frac{20\,V}{16.03\,mA} = 1.247\,k\Omega$; *Standard: 1k2*
- $P_{Ro1} = R_{o1} \times I_{o2}^2 = 1200\,\Omega \times 0.01603\,A^2 = 0.308\,W$ *(Choice 5W, safety reason)*

- $I_{03} = 0.68mA + 2 \times 6mA = 12.68\ mA$
- $U_{R02} = U_2 - U_3 = 410\ V - 300\ V = 110\ V$
- $R_{02} = \frac{U_{Ro2}}{I_{o3}} = \frac{110\,V}{12.68\,mA} = 8.67\,k\Omega$; *Standard: 8k2*
- $P_{Ro2} = R_{o2} \times I_{o3}^2 = 8200\,\Omega \times 0.01268\,A^2 = 1.318\,W$ *(Choice: 7 W; safety reason)*

#Note
To shorten the calculation process, the experience from the "Power Supply" Chapter (Detailed calculation of this type of high voltage power supply) can be used, as well as the experience from the calculation of the power supply of the SE amplifiers discussed in the text above.

Heater power supply
Input and driver tubes: DC type
ECC82: U_f = 12.6 V, 0.15 A
ECC83: U_f = 12.6 V, 0.15 A
A voltage regulator using the LM317 can be used.
If a common heater power supply is used for both channels it is necessary to use: voltage regulator 12.6 V / 0.6 A.

Output tubes: AC type
KT 88: 6.3 V / 1.6 A
Since a common power transformer is used to power both channels:
4 × 6.3 V / 1.6 A (Secondary of power transformer).

Power transformer:
Secondary:

- U_{S1} = 315 V / 0.4 A (high voltage); P_{S1} = 126W
- U_{S2} = 6.3 V / 1.6 A (KT88); P_{S2} = 10 W
- U_{S3} = 6.3 V / 1.6 A (KT88); P_{S3} = 10 W
- U_{S4} = 6.3 V / 1.6 A (KT88); P_{S4} = 10 W
- U_{S5} = 6.3 V / 1.6 A (KT88); P_{S5} = 10 W
- U_{S6} = 13 V / 1 A (heater voltage regulator, input and driver tubes); P_{S6} = 13 W.

Power transformer, total power of secondary:
$P_S = P_{S1} + P_{S2} + P_{S3} + P_{S4} + P_{S5} + P_{S6} = 179\ W$

In order to reduce the operating temperature of the transformer (heating of the secondary windings caused by heat dissipation at the DC resistance of the windings) it is practiced to increase the cross section of the wire - increase the load current of the secondary:

Secondary winding:

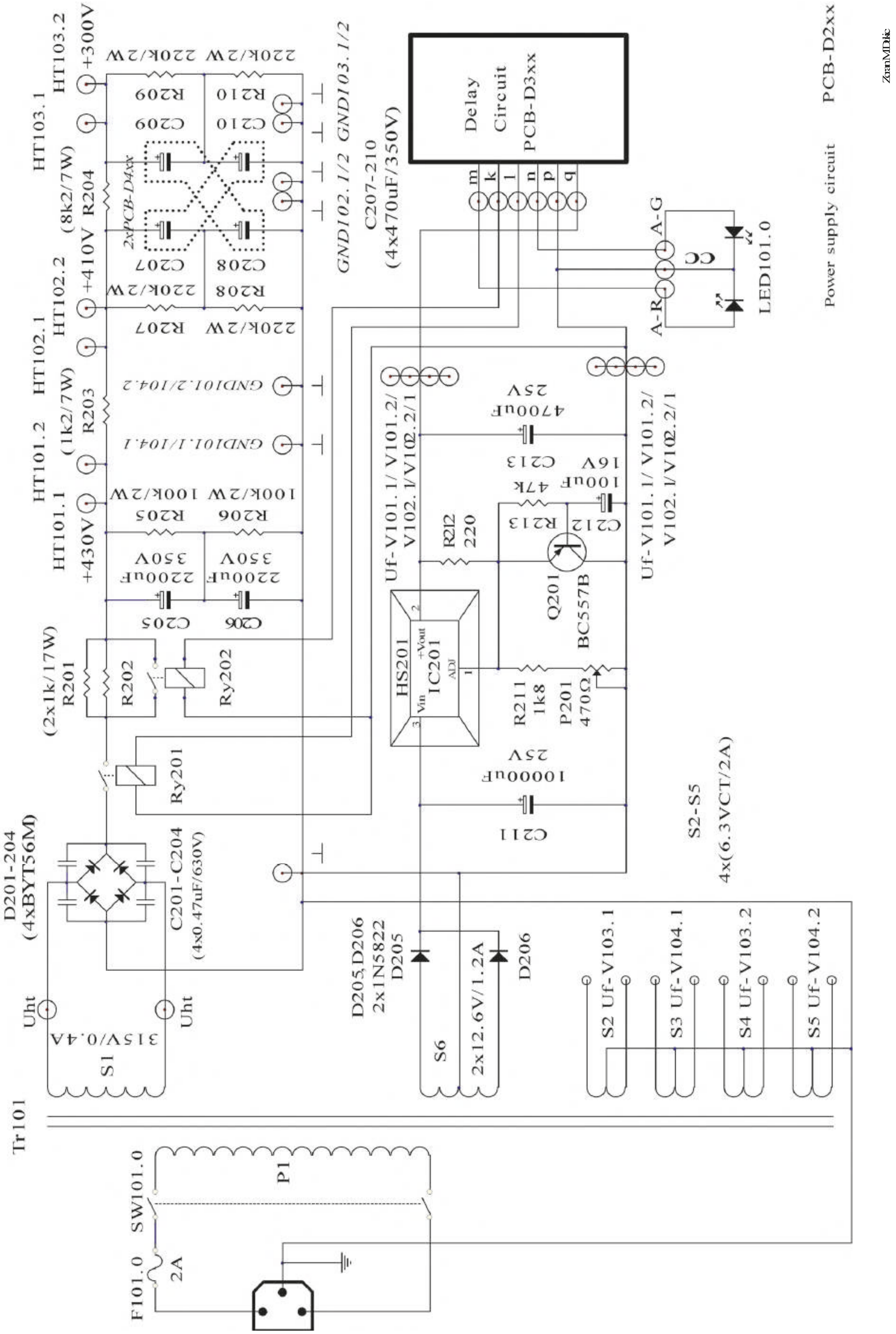
- U_{S1} = *315 V / 0.5 A (high voltage);* P_{S1} = *157.5 W*
- U_{S2} = *6.3 V / 2 A (KT88);* P_{S2} = *12.6 W*
- U_{S3} = *6.3 V / 2 A (KT88);* P_{S3} = *12.6 W*
- U_{S4} = *6.3 V / 2 A (KT88);* P_{S4} = *12.6 W*
- U_{S5} = *6.3 V / 2 A (KT88);* P_{S5} = *12.6 W*
- U_{S6} = *13 V / 1.2 A (heater voltage regulator, input and drivers tubes);* P_{S6} = *15.6 W.*

Total power of secondary:
$P_S = P_{S1} + P_{S2} + P_{S3} + P_{S4} + P_{S5} + P_{S6} = 223.5\ W$

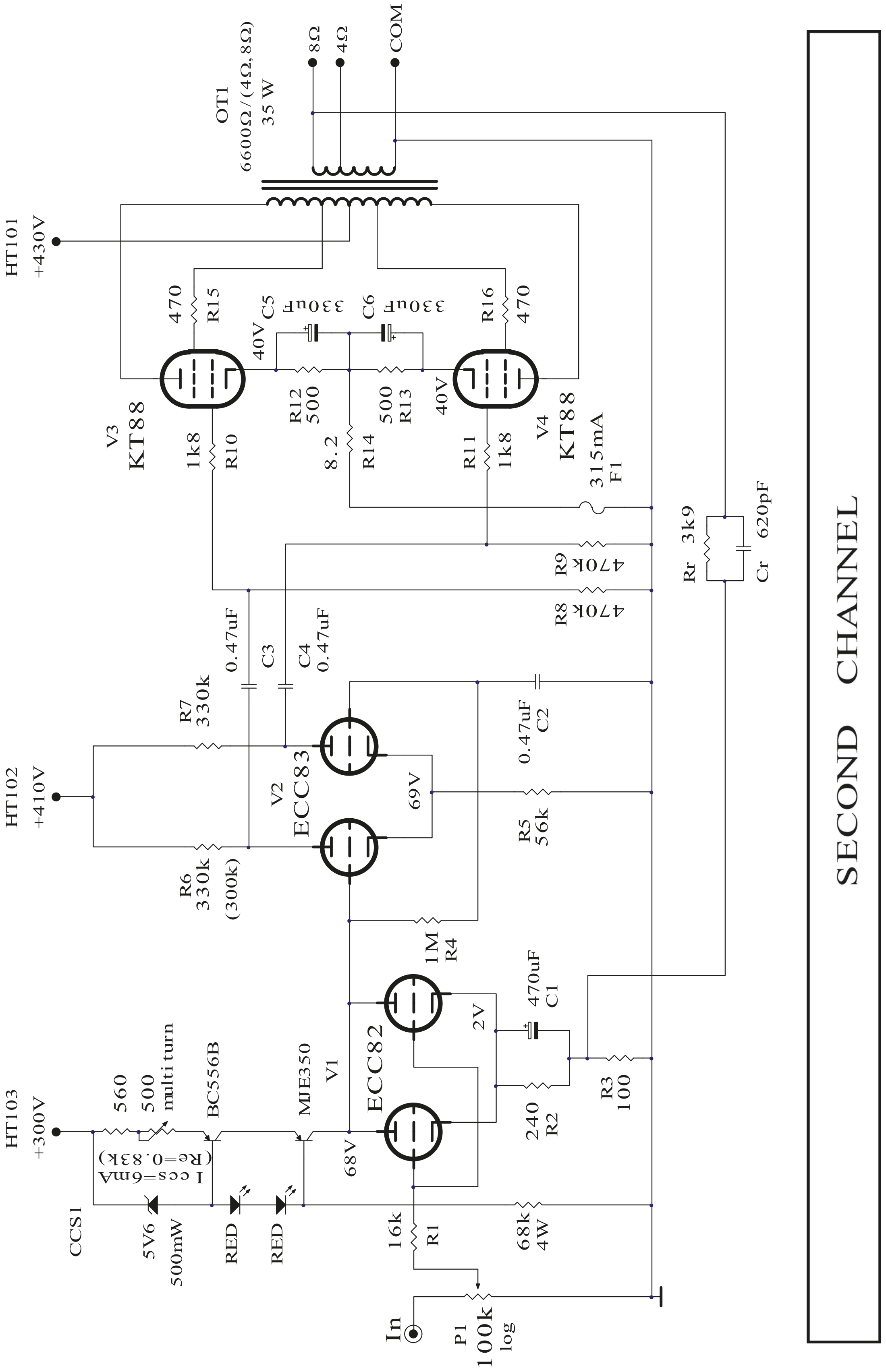
Power of the primary:
$P_P = 1.5 \times P_S = 1.5 \times 223.5\ W = 335.25\ W$ ($U_P = 230\ V$, $I_P = 1.457\ A$)

Parts selection:
The same criteria as for the selection of the components of the SE amplifier described above.

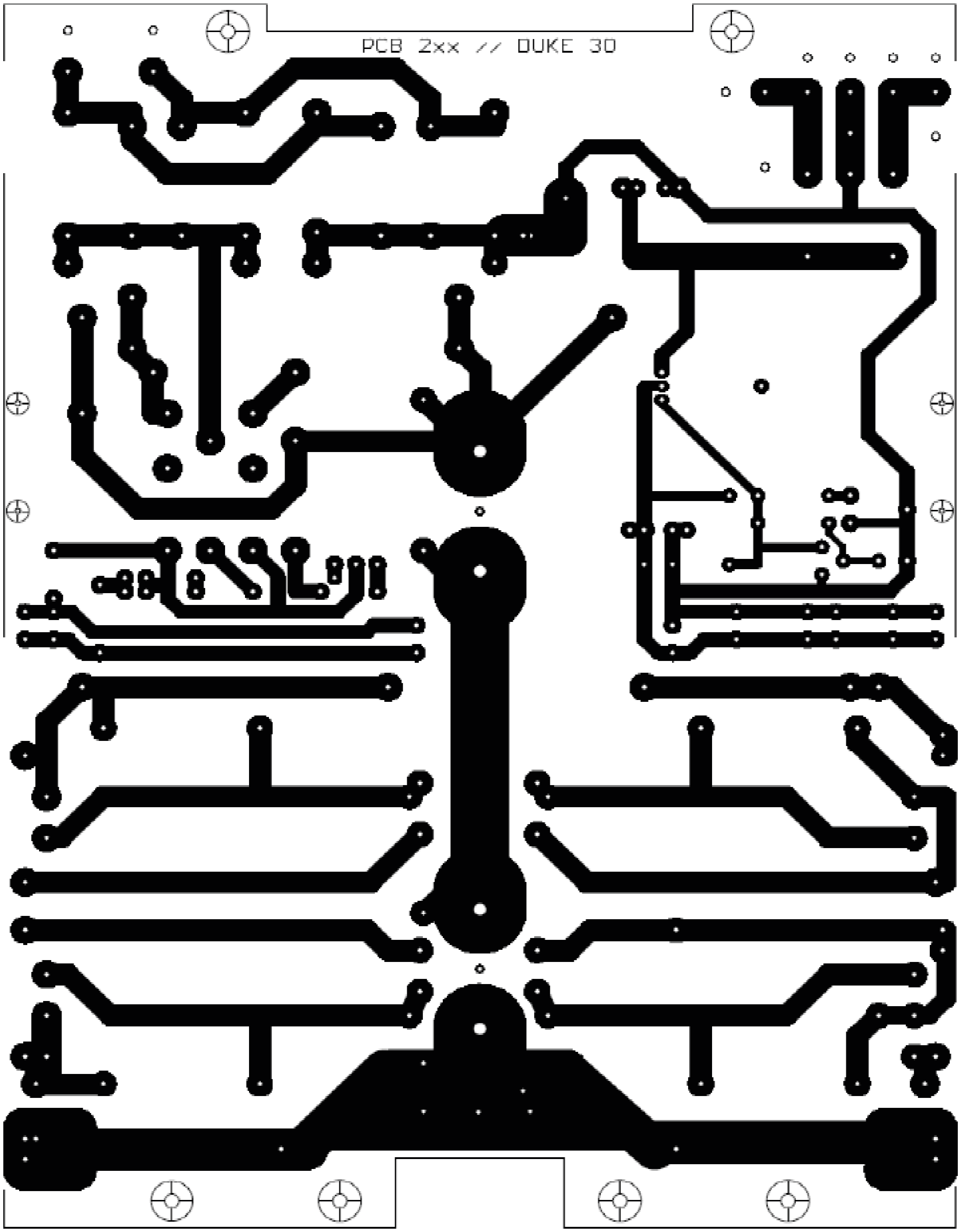
Power Supply, electric diagram



Power Amplifier, electric diagram

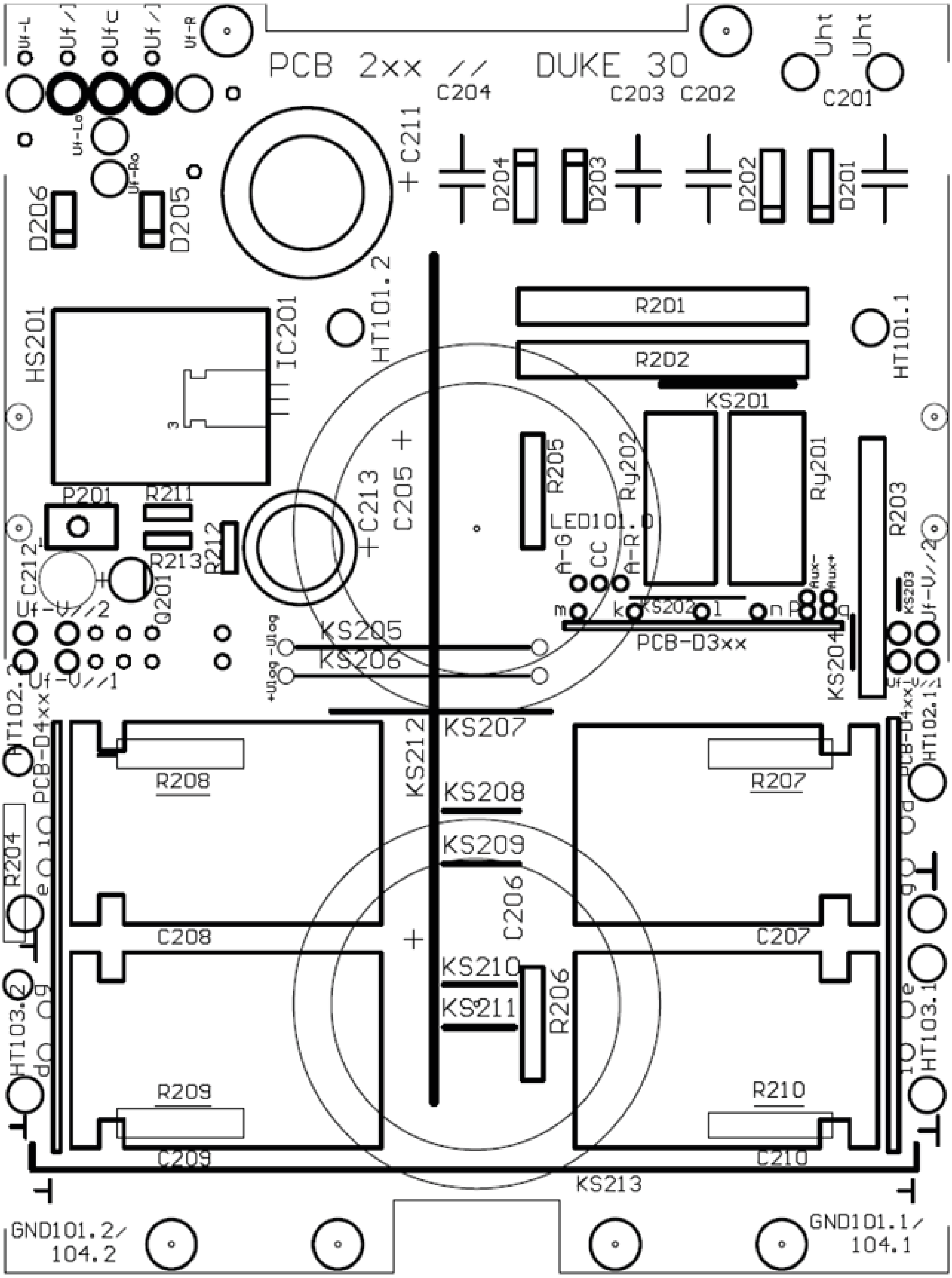


Power supply, PCB



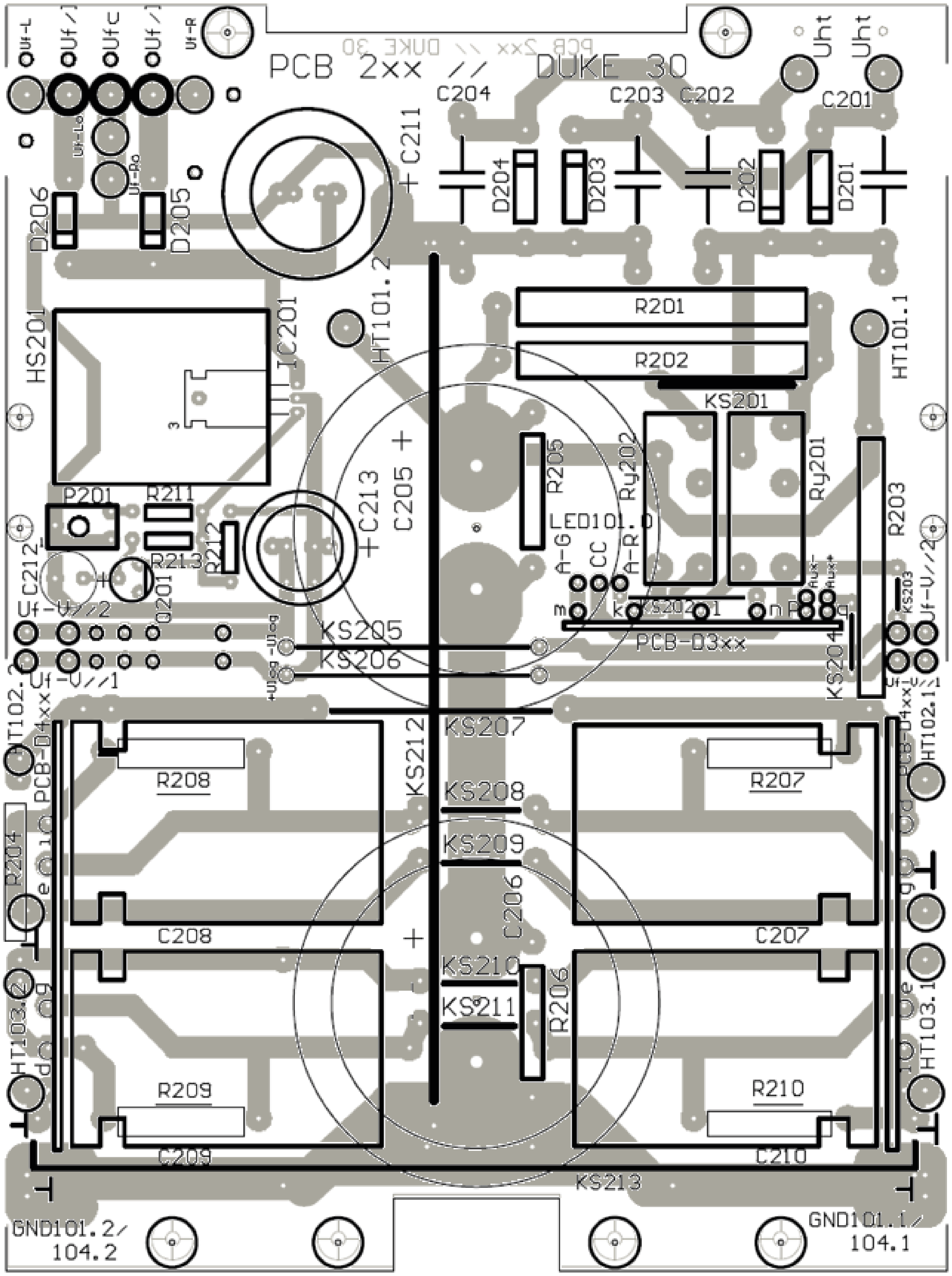
Cu side

Top overlay – components layout



Power supply PCB

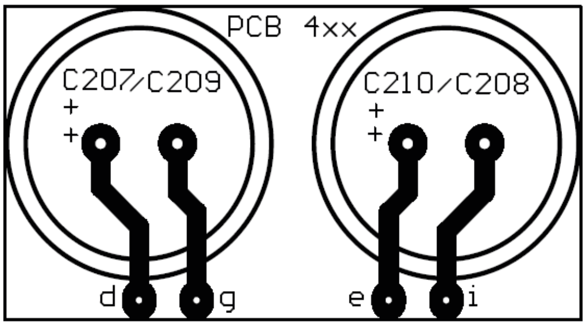
Composite



As can be seen in the power supply PCB component layout drawing, three smaller PCBs are mounted on the power supply mother board:

1. *PCB-D3xx – Previously described delay circuit.*

2. PCB 4xx – Power supply filter capacitors (C207 / C209, C208 / C210)



Why in this example, auxiliary PCBs with filter capacitors were used instead of filter capacitors mounted directly on the PCB motherboard of the power supply?
Answer: one of the unmentioned project requirements which, by the way, is not related to the performance of the amplifier - the height of the chassis must be small – 55 mm. Since the height of the housing of electrolytic capacitors (high voltage and high capacity) is larger than the diameter, usually, (in this example H = 50 mm and D = 35 mm), the only way to mount filter capacitors inside a chassis of given dimensions was to mount the capacitors horizontally. Auxiliary PCBs are mounted on the power supply mother board using the PCB edge pins (solid Cu wire). This is also an example of how sometimes specific project requirements are in conflict with optimal design solutions — a compromise solution.
However, such "illogical" requirements should not be ignored, because the reasons can be numerous and useful, such as the external look of the amplifier, minimizing the material used to make the chassis, available processing technologies, ..

 # General notes

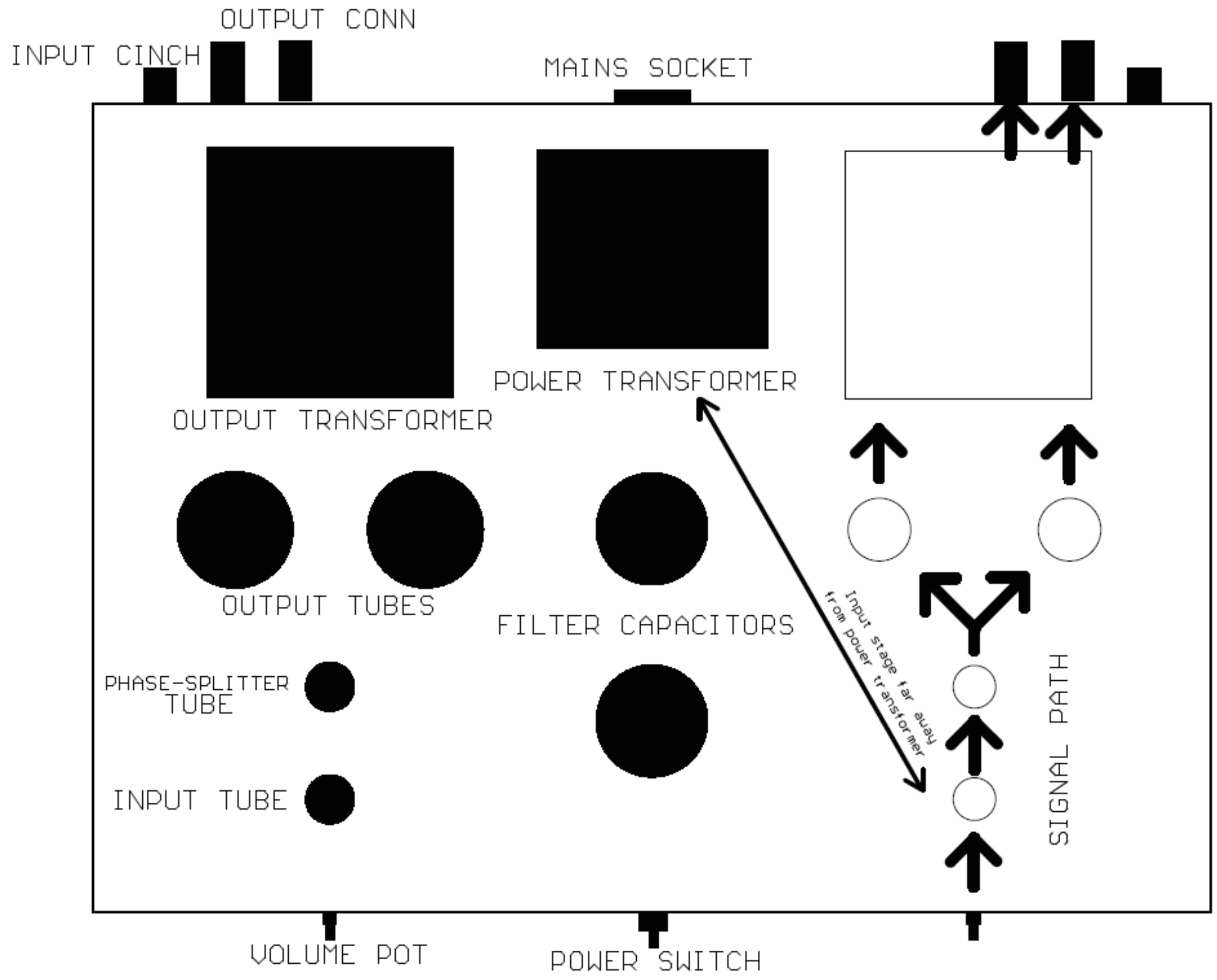
- *Before starting the construction of mechanical parts of the amplifier (chassis, transformer shield, ...) it is necessary to collect all electronic, electrical and electromechanical components or their detailed technical data (not only electrical data and electrical characteristics but also detailed information about their physical dimensions and shapes and even the masses).*
- *When constructing the chassis (dimensions and shape of the chassis), take into account not only the possibility of physical mounting and carrying of electromechanical and electronic components, but also the construction of the chassis which enables suitable arrangement of components (layout) in accordance with the "golden rules of designing".*
- *The development and construction of the chassis begins with the development of the amplifier parts layout concept (taking care about "golden rules of designing").*

- ***"Golden rules of designing" – basics***

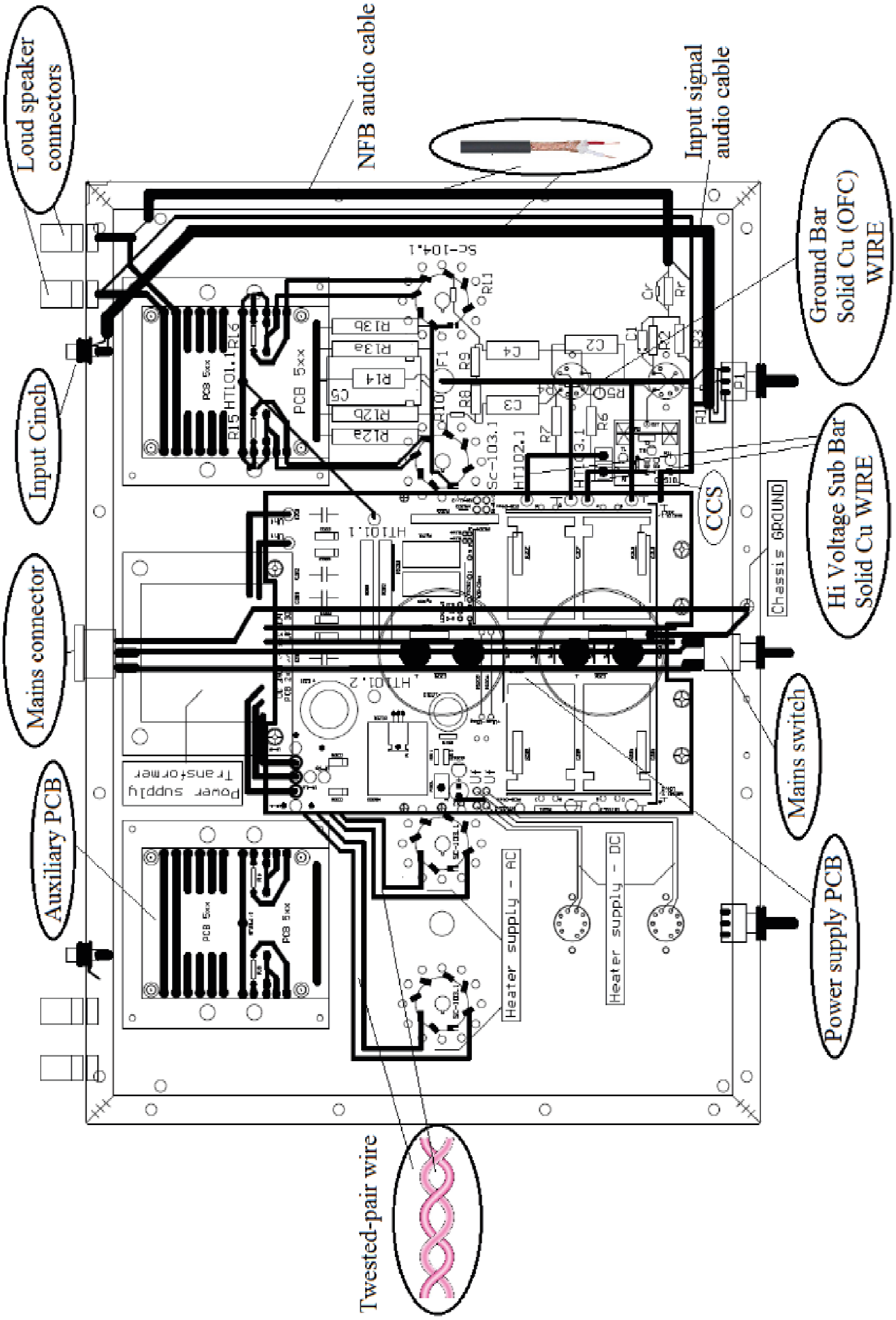
  1. ***The shortest possible signal path.***
  2. ***The input stage is located as far as possible from the power transformer.***

3. ***Location, distance and orientation of transformers (and other sources of magnetic and electromagnetic fields), both relative to each other and relative to the location of the components in the path of the audio signal, must be designed in such way to ensure minimal unwanted electromagnetic effects on each other.***
4. ***Locations of components that radiate thermal energy must be designed in such way to ensure minimal thermal effects to the other components of the amplifier.***
5. ***The chassis must be a mechanically stable structure capable of carrying heavy electrical and electromechanical parts.***
6. ***The chassis must provide good heat dissipation of all amplifier components and air circulation - ventilation.***
7. ***Constructively, the chassis should meet the ergonomic rules - easy, logical and obvious manipulation of amplifier commands (volume control, power supply switch, fuses, indications…) and their location. Also, the chassis design must provide easy connection of the amplifier to external equipment (mains connector, input connectors, and loudspeaker connectors,…).***
8. ***Looks — attractive***

Chassis – concept

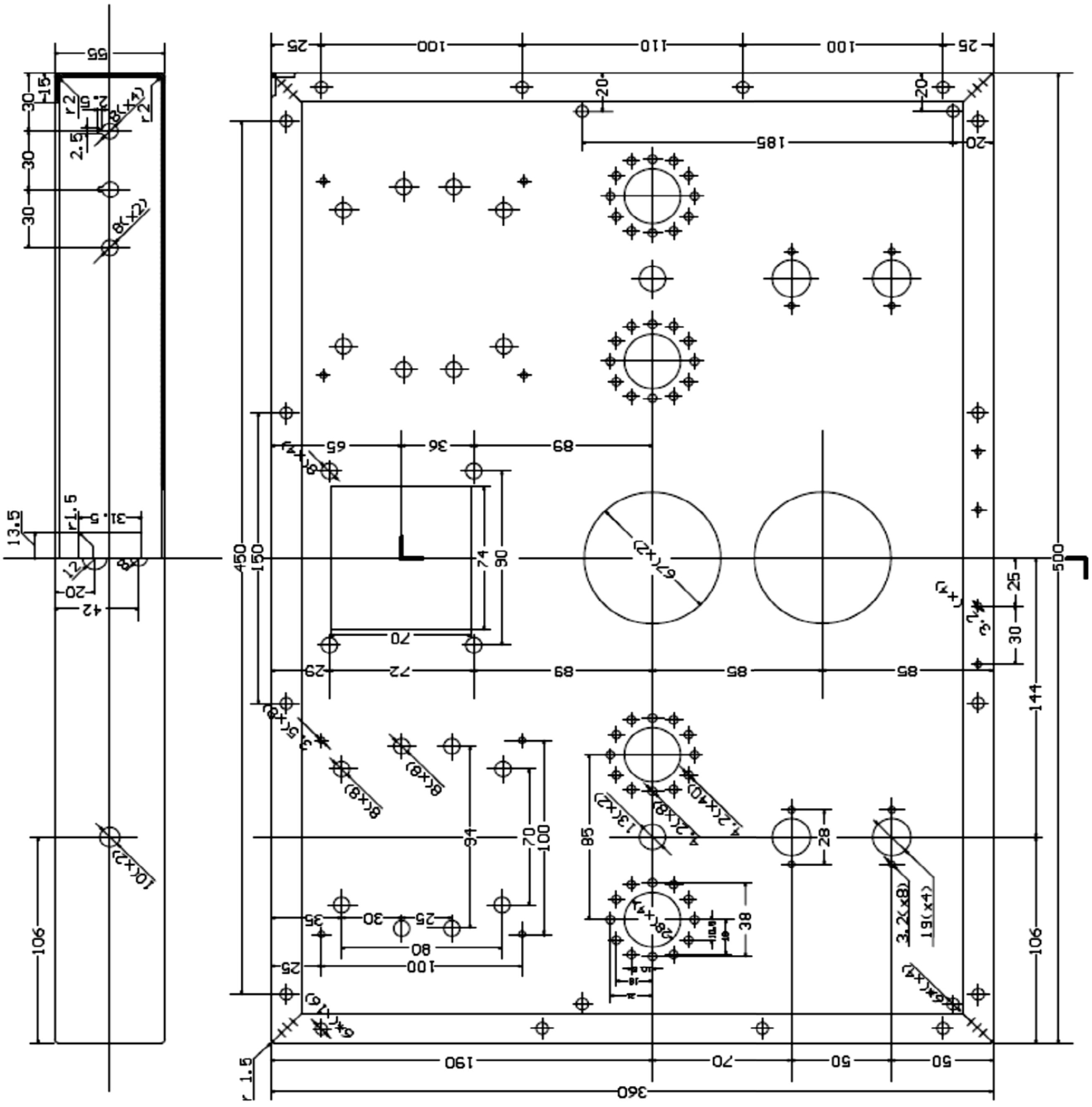


Layout of components inside the chassis – concept



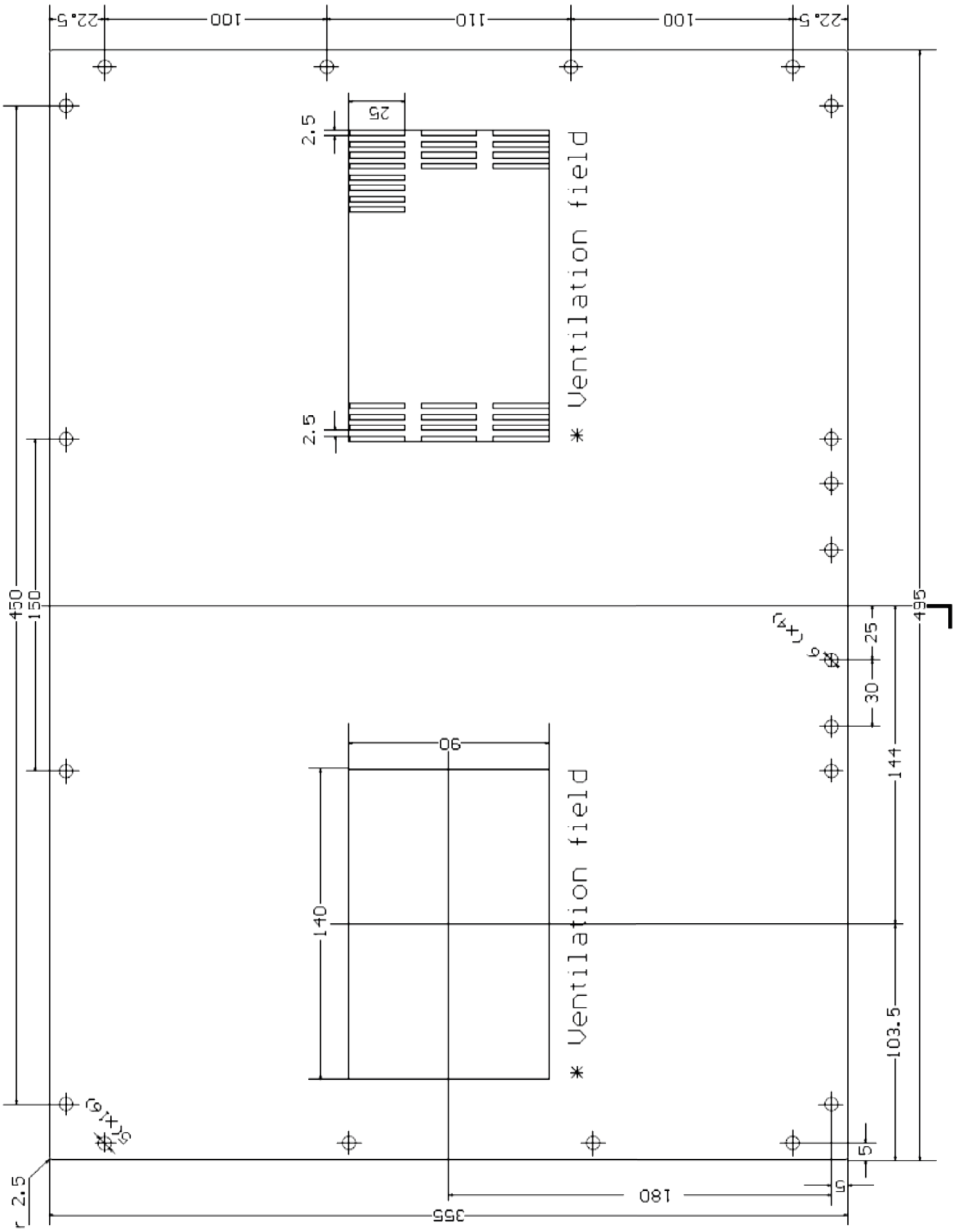
Chassis drawing

Dimensions: 500 × 360 × 55 mm



Material: Stainless steel # 1.2 mm

Chassis bottom cover



Dimension: (495 × 355) mm

Material: Stainless steel # 1.2 mm

 # Note 1

Other shapes of ventilation field openings can be used (for example holes of appropriate Ø).
For safety reasons, the holes (opening width) must not be larger than 8 mm.
In any case, the ventilation field must provide sufficient airflow to cool the inside of the chassis.

 # Note 2
The chassis feet:

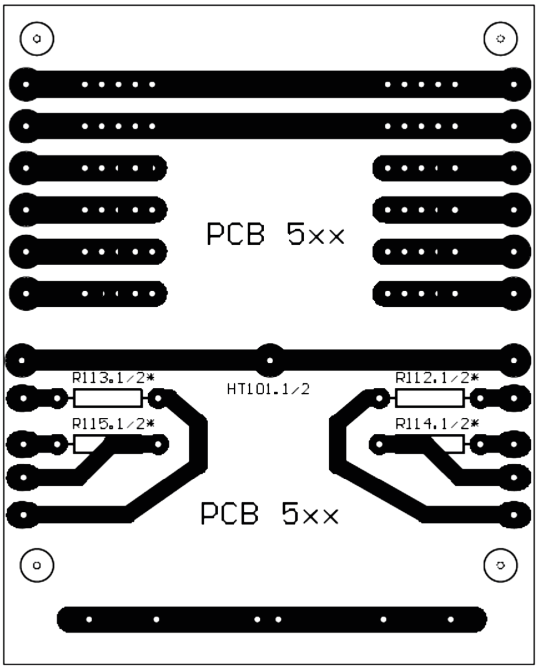
Although they seem insignificant, the feet of the amplifier have a significant role in, and effects on, the operation of the amplifier.

It is good to use chassis feet with rubber (or other shock-absorbing material) on the bottom side of the foot in order to reduce the transmission of mechanical vibration from the substrate to the amplifier and vice versa.

Also, the chassis feet should be of a certain height to ensure air circulation and adequate cooling of the amplifier.

Some details of the layout of components inside the chassis – concept

- *Auxiliary PCB 5xx*



The wire leads of the output transformer coils are connected to other parts of the amplifier using an auxiliary PCB 5xx. The use of the auxiliary PCB 5xx facilitates the combination and connection of transformer windings to determine the output impedance of the amplifier (4 Ω, 8 Ω) by simply connecting a jumpers between the corresponding Cu traces of the PCB.

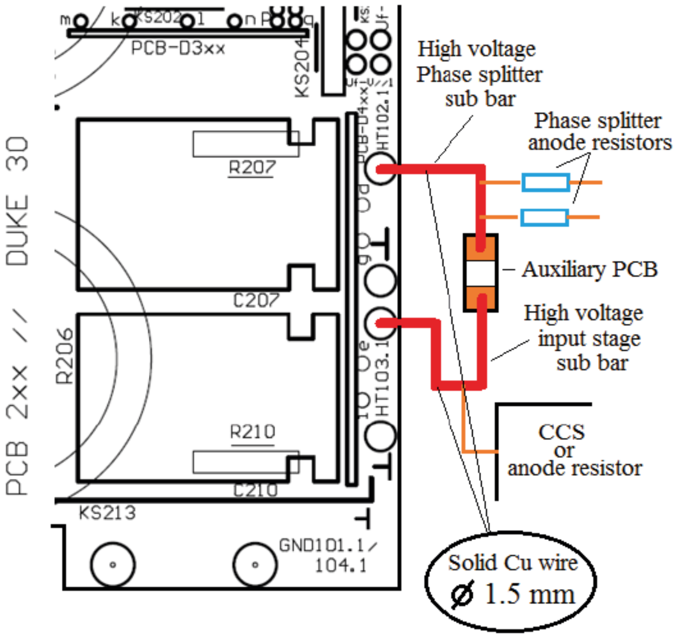
Another function of the PCB 5xx may be to carry some other parts of the amplifier such as the cathode resistors of the output tubes which can sometimes be large in size (the case of PP amplifier in this example).

The reason for mentioning this detail of the amplifier construction is this:

In order to facilitate the technology of making amplifiers, various auxiliary constructive solutions can be used (depending on the skills, ideas, imagination and experience of the amplifier designer).

Another example:

- *Auxiliary bar for high voltage supply of phase splitter and input stage*



Solid Ø1.5 mm Cu wires bent into the appropriate shape and soldered to the mother board of the power supply (points HT 102 and HT 103) and to a small auxiliary PCB make a very stable mechanical and electrical support structure for the anode resistors of the phase splitter and input stage.

Power Amplifier - Assembly procedure

- ***First step***
  Installation of mechanical, electromechanical and electronic components (mains socket, mains switch, fuse holders, input RCA (cinch) sockets, loudspeakers terminal connectors, tube sockets, volume potentiometers, rubber grommets, power supply transformer, output transformers, transformer shields, spacers, some mechanical holders,...) to the chassis. Use screws of appropriate diameter and length, nuts and washers.

- ***Second step***
  Installation of already ready-made (completed) power supply PCB (mounted and soldered all electronic components including auxiliary wire bars and wiring wires) and auxiliary PCBs such as PCB 5xx. Power transformer mounting screws, some spacers and special holders and supports can be used to mount the power supply PCB, for example. Also, output transformers (or transformer shields) mounting screws can be used to mount the auxiliary PCB 5xx, for example.

- ***Third step***
  Wiring and soldering.
  - *Mount the ground bar on particular supports or use some of the already mounted parts of the amplifier as supports. Solder it to the fuse holder of the cathode circuit of the output tubes or to the central pins of the tube sockets or to the power supply PCB or to the volume potentiometer, for example.*
  - *Wire the mains socket, mains fuse, mains switch, and power transformer. Wiring must be done very carefully using high quality insulated wires and required cross-sections (it would be best to do this using a shielded power cable). Special attention should be paid to chassis grounding.*
  - *Wiring of tube heaters. Wiring of output tube heaters: twisted pair insulated stranded wires of required cross-section. Wiring of input and phase – splitter tube heaters: If a DC power supply is used to supply the input and phase – splitter tube heaters, as is the case in this example, it is not necessary to use twisted pair wires. Also, it is not necessary to use stranded wires — solid insulated wires can be used (but it is still recommended to use a twisted pair wires).*
  - *It is recommended to place the wires for wiring the tube heaters close to the chassis surface.*

- *Install and solder the audio cables to connect the input RCA (cinch) sockets and volume potentiometers. Install and solder NFB audio cables. Audio cables and ground wire place close to the chassis surface (base surface or wall surfaces). And, very importantly and necessarily, solder the ground wire to the chassis.*
- *Install and solder other wires (output transformer secondary windings leads – connect to loudspeaker terminal connectors or to PCB 5xx, output transformer secondary windings leads — connect to anode and screen grid pins of output tube sockets or to PCB 5xx, central tap lead of the output transformer primary winding — connect to power supply PCB (high voltage supply) or to PCB 5xx, solder the jumpers for selection the output impedance of the output transformer to the PCB 5xx,…).*
- *Install and solder all other electronic components: resistors, capacitors, electrolytic capacitors, ready-made CCS, some wires, …*

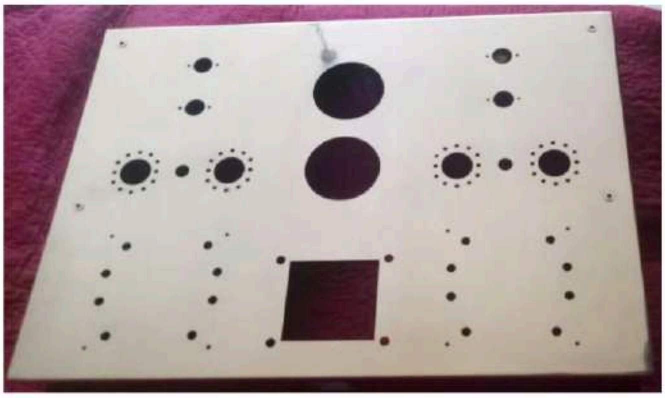
 # Note
All electronics component leads, uninsulated wire ends, tube socket pins, transformer pins and leads, connector pins, ground bar, must be clean and degreased. Use rosin alcohol tincture to clean the contact surfaces, if it is necessary – never use aggressive acid paste.

Use high quality Tin Lead Rosin Core Solder Wire:

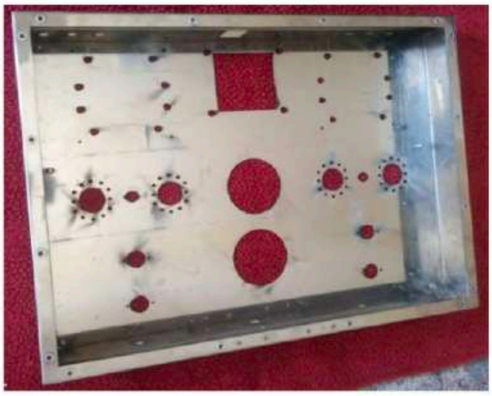
(Commonly used: Sn 60% - Pb 40%, Flux 2%, Ø 0.6 mm, 0.8 mm, 1 mm).

Use a quality 35 to 75 W soldering iron with adjustable temperature control from 180 to 450 ^{0}C.
It is common to use an operating temperature between 320 ^{0}C and 380 ^{0}C.

Chassis



Top



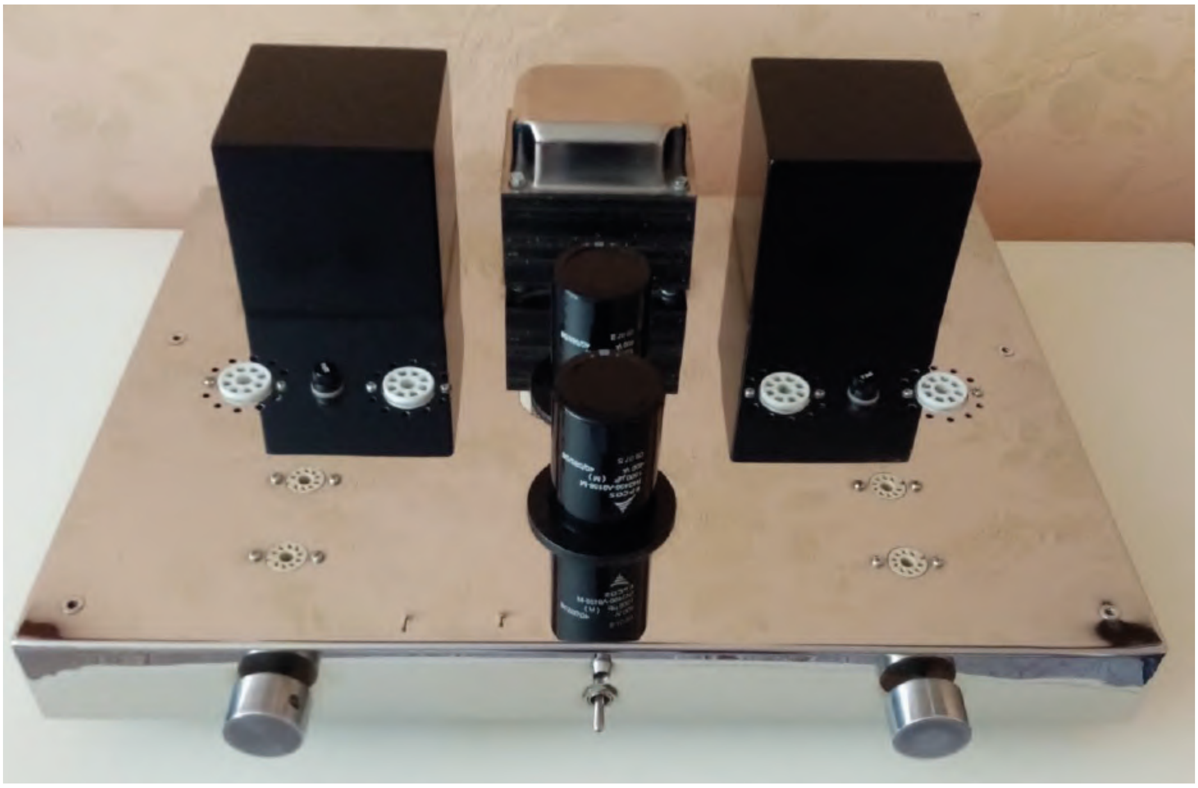
Bottom

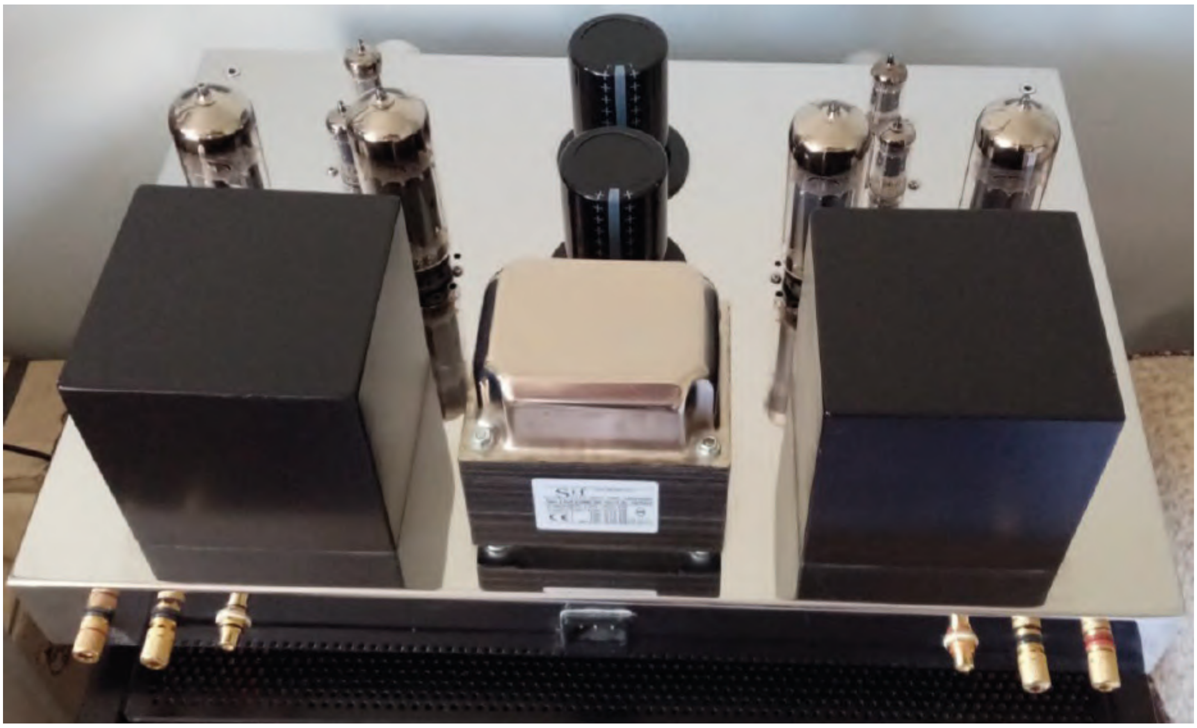
Corner welding

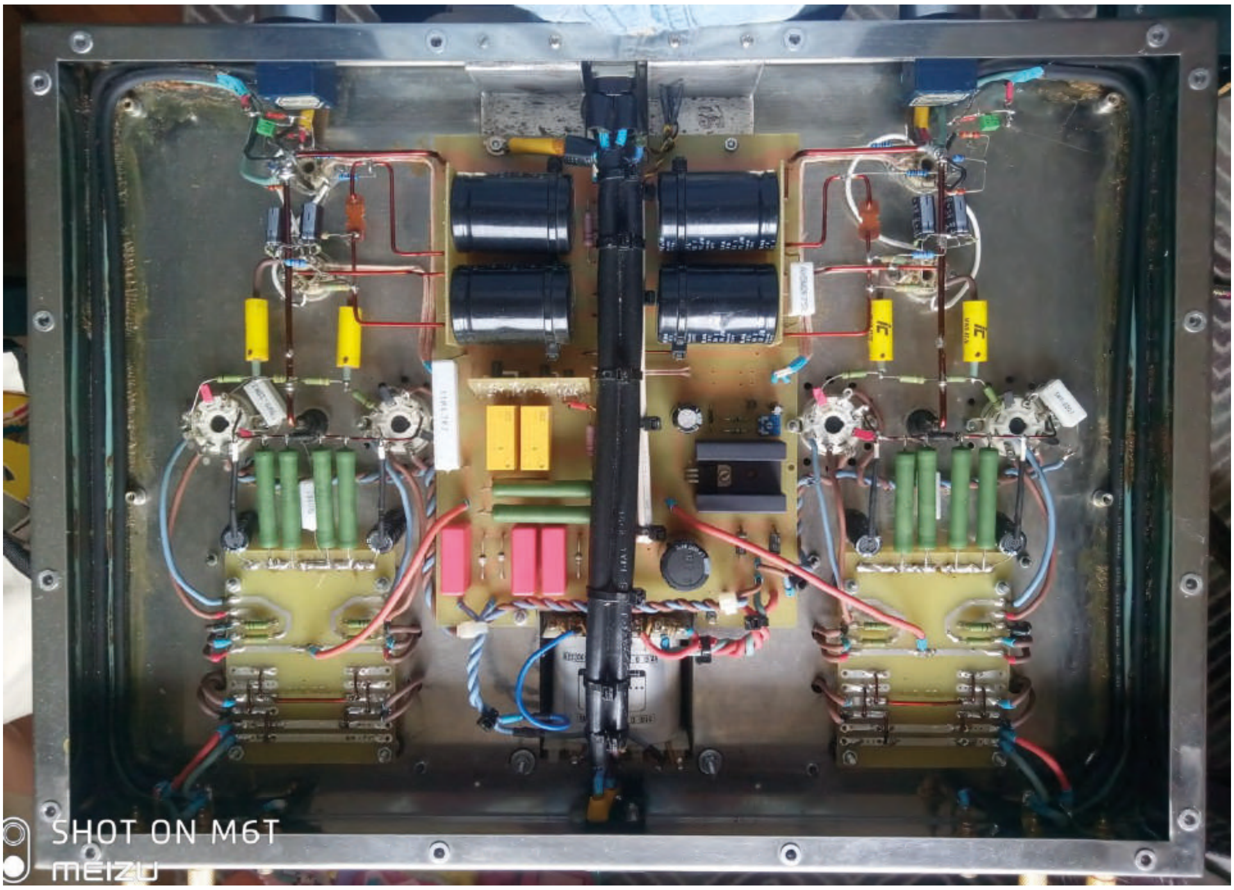
Inside

Outside – polished

Bottom cover: ventilation field & chassis feet

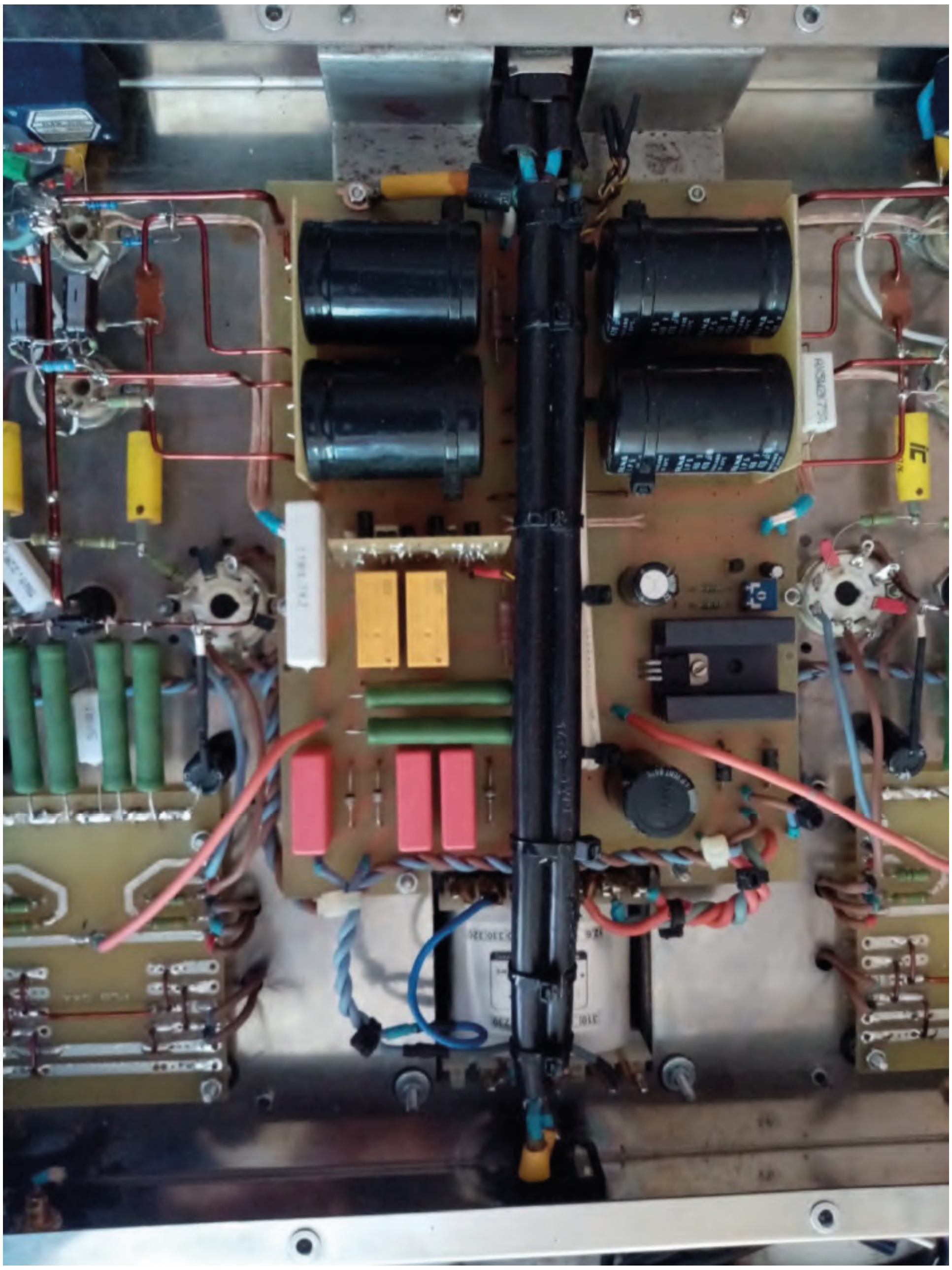


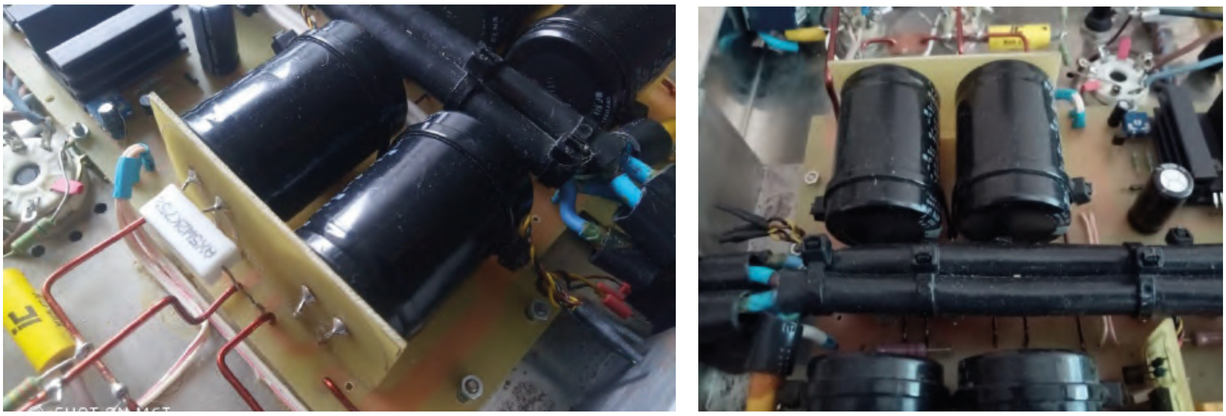




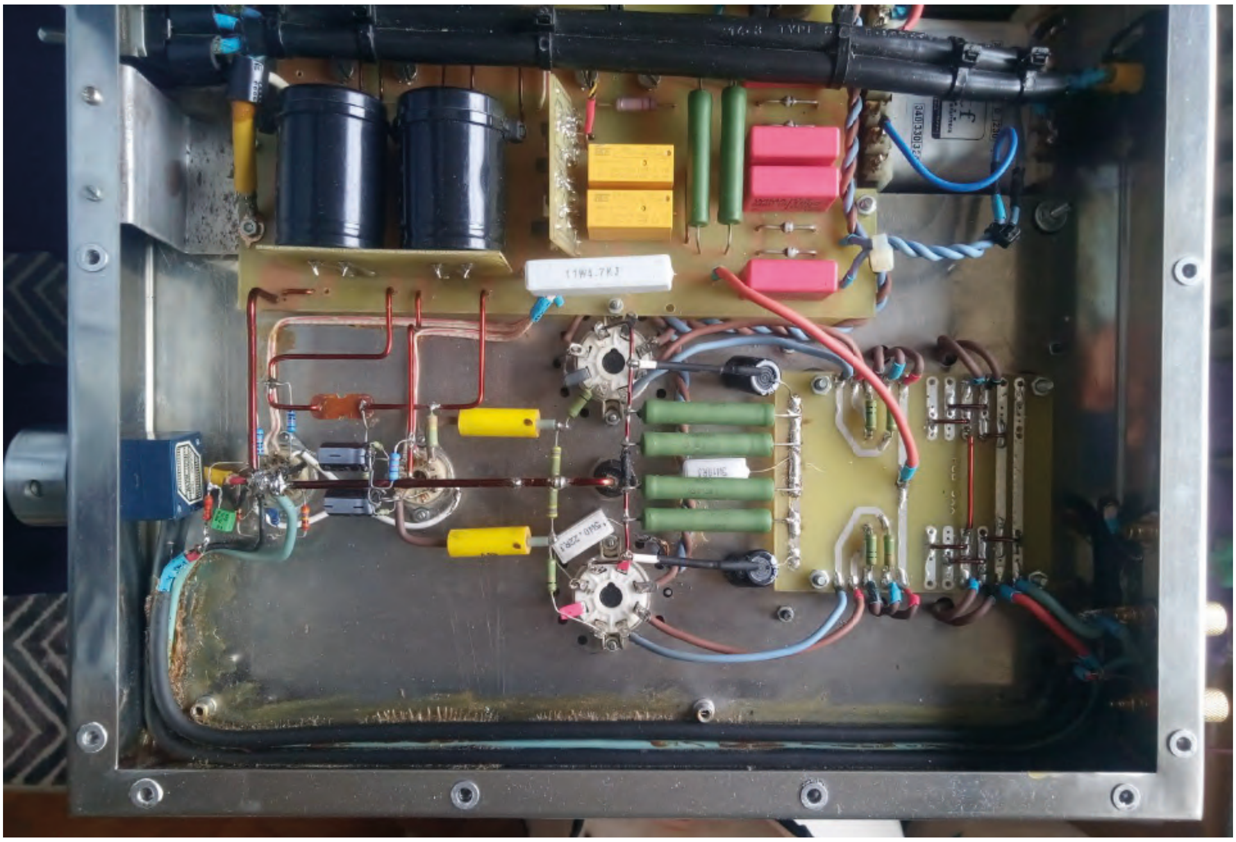
Chassis interior

Common power supply for both channels

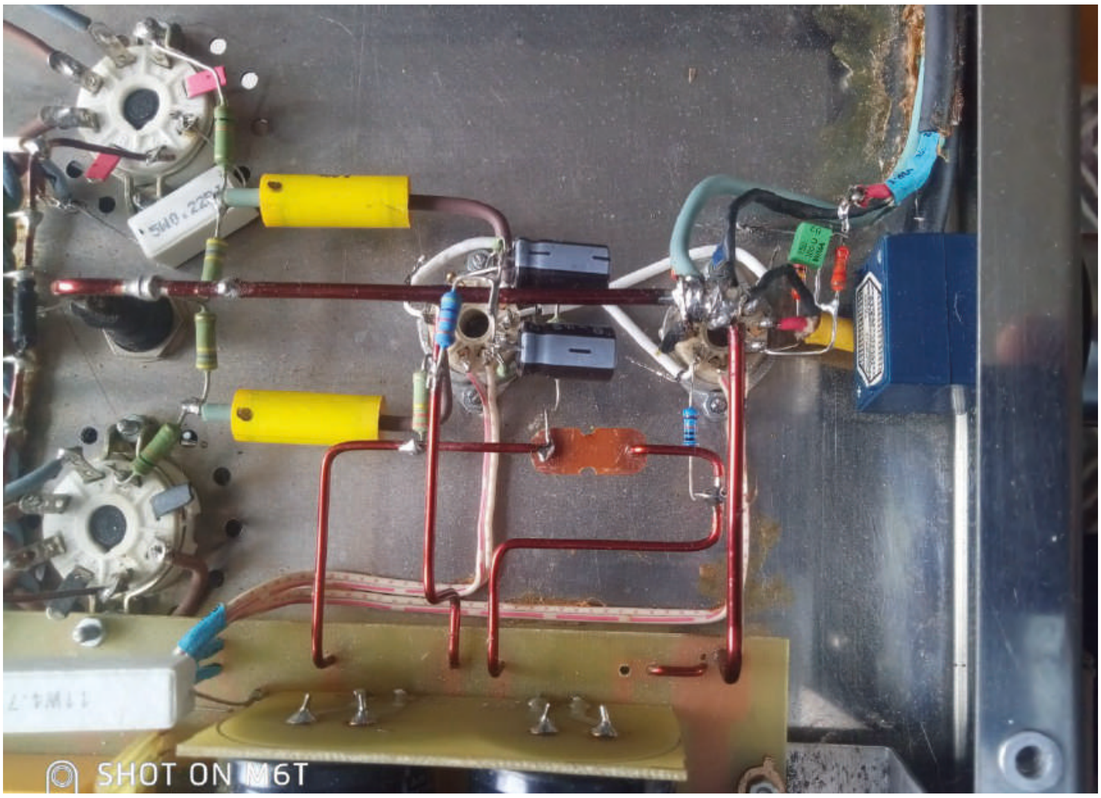




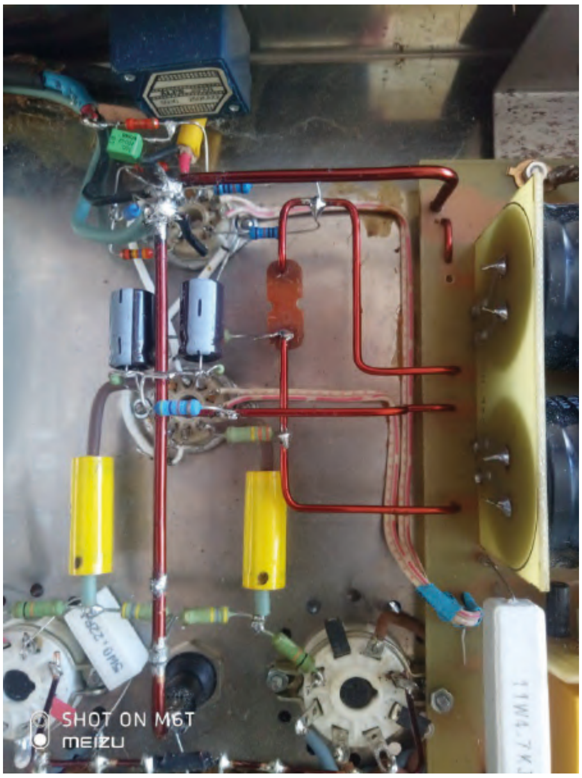
Auxiliary PCB - horizontal mounting of capacitors



One channel

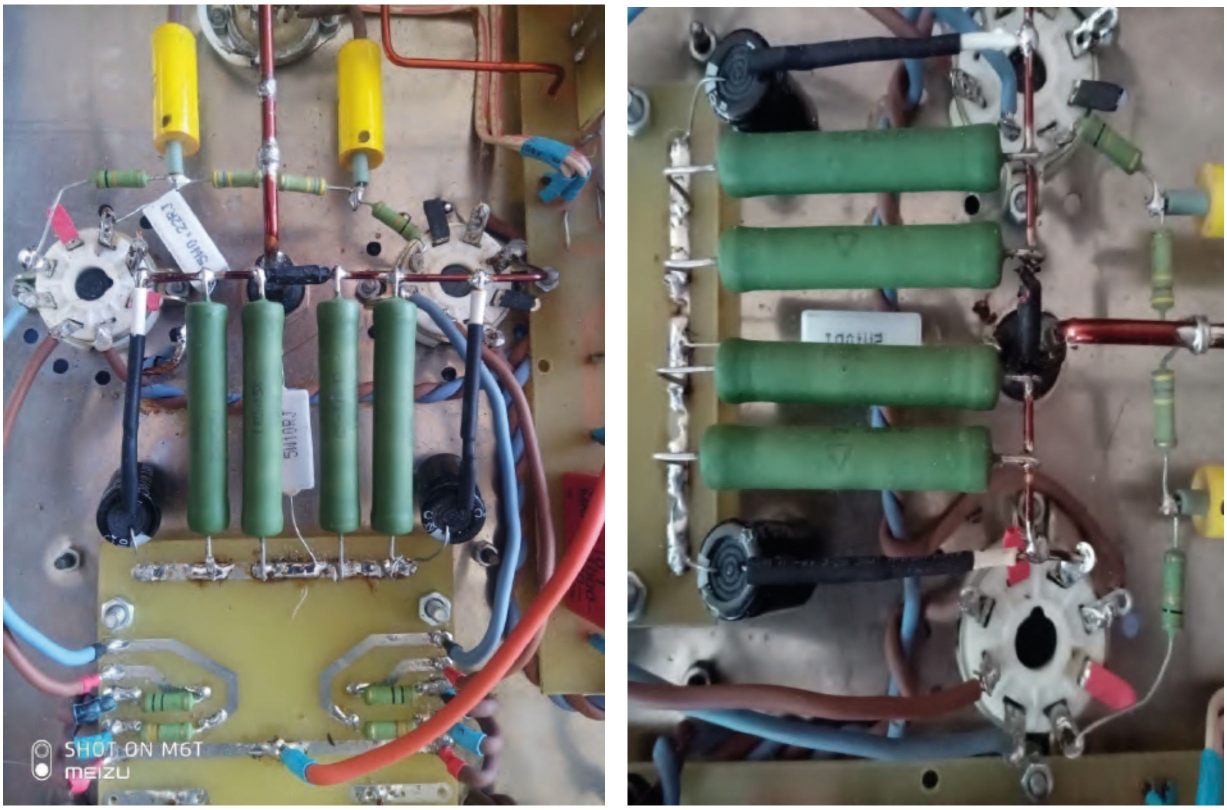


Short signal path



Main ground bar

Auxiliary bars



Auxiliary PCBs and cathode resistors of the output tubes

It would not be fair or correct to end this chapter without mentioning some significant constructions of amplifiers that in certain periods of time determined the direction of development of audio amplifier design.

Williamson amplifier

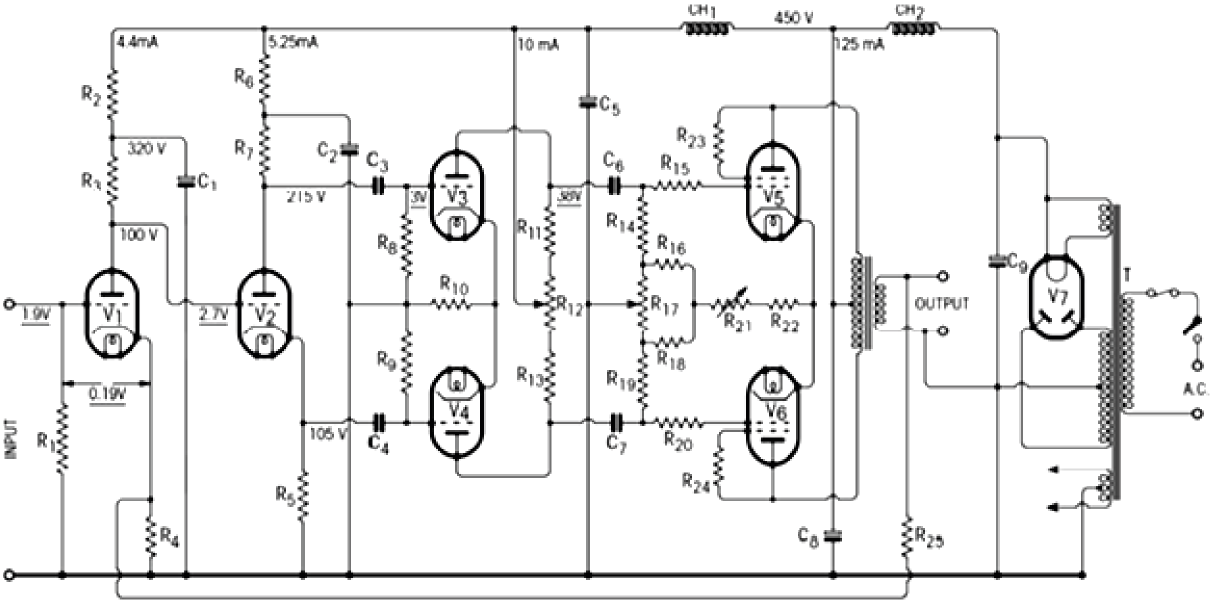
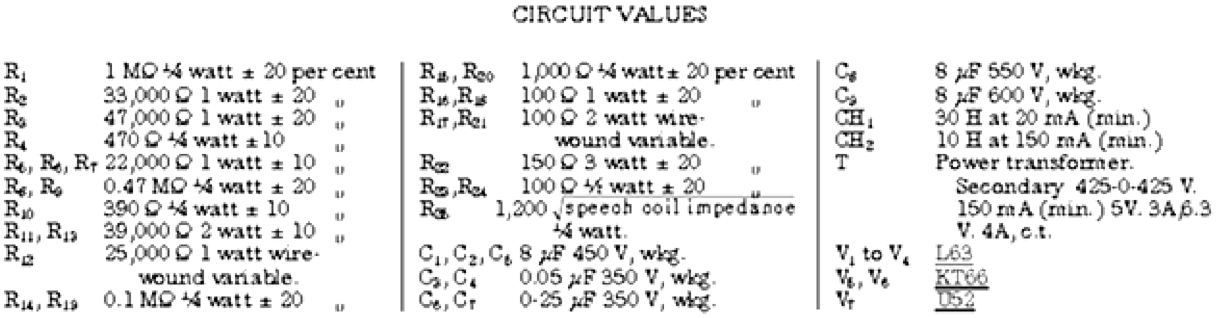


Fig. 5. Circuit diagram of complete amplifier. Voltages underlined are peak signal voltages at 15 watts output.



CIRCUIT VALUES

R_1	1 MΩ ¼ watt ± 20 per cent
R_2	33,000 Ω 1 watt ± 20 ,,
R_3	47,000 Ω 1 watt ± 20 ,,
R_4	470 Ω ¼ watt ± 10 ,,
R_5, R_6, R_7	22,000 Ω 1 watt ± 10 ,,
R_8, R_9	0.47 MΩ ¼ watt ± 20 ,,
R_{10}	390 Ω ¼ watt ± 10 ,,
R_{11}, R_{13}	39,000 Ω 2 watt ± 10 ,,
R_{12}	25,000 Ω 1 watt wire-wound variable.
R_{14}, R_{19}	0.1 MΩ ¼ watt ± 20 ,,
R_{15}, R_{20}	1,000 Ω ¼ watt ± 20 per cent
R_{16}, R_{18}	100 Ω 1 watt ± 20 ,,
R_{17}, R_{21}	100 Ω 2 watt wire-wound variable.
R_{22}	150 Ω 3 watt ± 20 ,,
R_{23}, R_{24}	100 Ω ¼ watt ± 20 ,,
R_{25}	1,200 $\sqrt{\text{speech coil impedance}}$ ¼ watt.
C_1, C_2, C_5	8 µF 450 V, wkg.
C_3, C_4	0.05 µF 350 V, wkg.
C_6, C_7	0·25 µF 350 V, wkg.
C_8	8 µF 550 V, wkg.
C_9	8 µF 600 V, wkg.
CH_1	30 H at 20 mA (min.)
CH_2	10 H at 150 mA (min.)
T	Power transformer. Secondary 425-0-425 V. 150 mA (min.) 5V. 3A, 6.3 V. 4A, c.t.
V_1 to V_4	L63
V_5, V_6	KT66
V_7	U52

Year 1947
Articles published in “Wireless World”, “Design for a High-quality Amplifier” by D.T.N. Williamson

In a series of articles, Mr. Williamson presented his opinion on what characteristics a high quality amplifier should have:

- Negligible non-linear distortion (0.1%) up to the maximum, rated output power and at all audible frequencies from 10 to 20000 Hz.
- Linear frequency response and constant output power at all audible frequencies.
- Negligible phase shift within the audible frequency range.
- Good transient response.
- Low output impedance (damping factor: 20 ÷ 30).
- Adequate power reserve (output power of 15–20 W for reproduction of orchestral music via a dynamic loudspeaker, or 10 W for a horn loudspeaker).

The significance of these articles is that the views of Mr. Williamson stated in them have become unofficial standards for designing Hi End amplifiers.

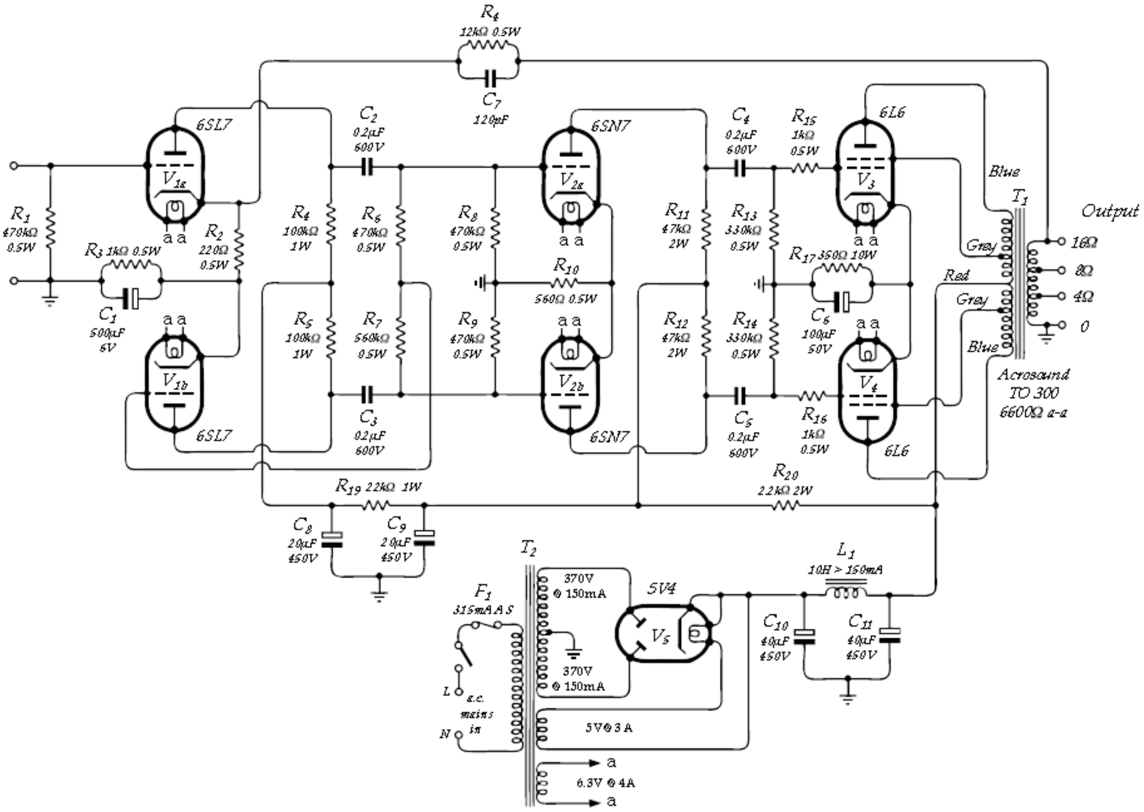
Williamson amplifier description:

- All triodes in a class A four stage amplifier.
- Input stage: grounded cathode with resistance in a cathode circuit.
- Input stage connection to the phase splitter stage: DC coupling.
- Phase splitter: concertina type phase splitter.
- Driver stage with AC balancing circuit.
- Output stage: Push pull, triodes (tetrodes connected in a triode mode) operate in class A. Automatic bias with common cathode resistors network (without bypass capacitors) and with a DC balancing circuit.
- Output transformer primary impedance 10 000Ω (turns ratio 76 : 1)
- NFB (applied from the amplifier output to the input stage).

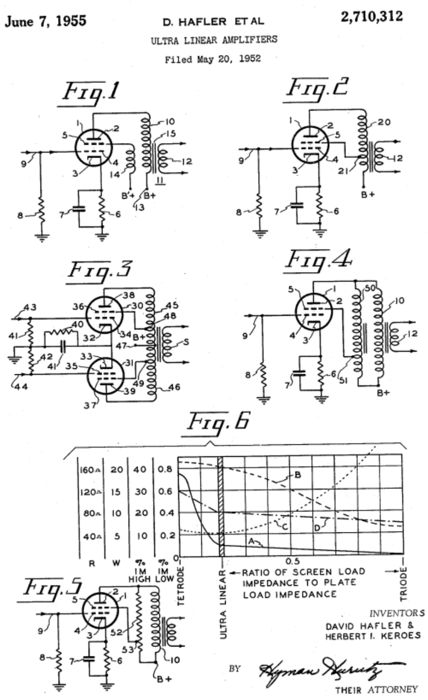
Mr. Williamson paid special attention to the construction of the output transformer:

“28A Super Silcor laminates 13/4 in stack. The winding consists of two identical interleaved coils each 11/2 in wide, wound on 11/4 in × 13/4 coil former. On each former is wound: 5 primary sections each consisting of 440 turns (5 layers, 88 turns per layer) of 30 S.W.G enameled copper wire interleaved with 2 mil. paper, alternating with 4 secondary sections, each consisting of 58 turns (2 layers, 29 turns per layer) of 19 s.w.g, enameled copper wire interleaved with 2 mil. paper. Each section is insulated from its neighbors by 3 layers of 5 mil. Empire tape.”

Year 1951



Audio Engineering, **“An Ultra–Linear Amplifier”** by David Hafler and Herbert I. Keroes



More power, other characteristic are almost the same as the triode push pull output.

Tubes: 6L6, 6SN7, 2 × 6L6
Output transformer:
Ultra linear push pull
Acrosound: Type TO - 300.
Primary: 6600 Ω plate to plate. (UL taps)

.

A story about designing an amplifier or a sad story about the life and end of a once very famous vacuum tube factory

Technical requirements:
Vacuum tube class A single ended power amplifier, (two mono blocks).
Output stage (classic design): two WE 300B connected in parallel, automatic bias, quiescent point recommended by Western Electric data sheet, anode output transformer - optimal load.
Driver stage: triode mode KT 88, anode load: CCS?
Input stage: triode WE 437A, anode load: CCS!

Power supply:
heater power supply: regulated DC power supply (all tubes).
output stage high voltage supply: highly filtered high voltage power supply with high capacity output filter capacitors.
driver and input stage high voltage supply: regulated high voltage power supply.

Comment:
It cannot be said that the design requirements are very common. The concept of amplifier (amplifier stage topology) and the choice of tubes (high cost WE 437A, use of power beam tetrode designed for use in output stage of an audio amplifier as a driver stage tube) are debatable. The design of the amplifier is pretentious in every aspect.

Why did the author of this book decide to design such an amplifier despite the indisputably reasonable above remarks?

The answer should be sought in the story of the life and end of the once well-known vacuum tube factory – Electronic industry Nis-Ei Radio Tube Factory.

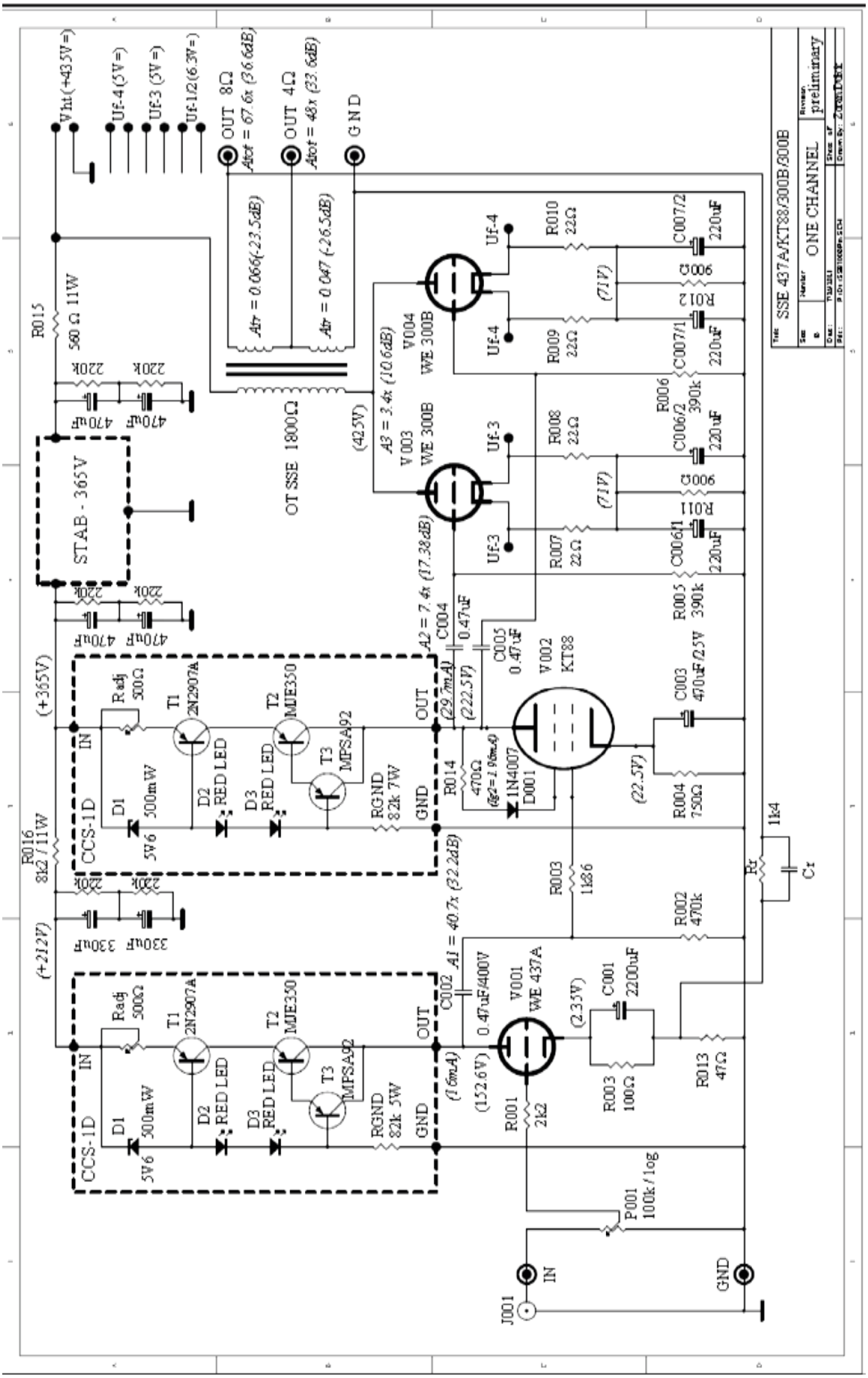
The story

The Ei Radio Tube Factory was established in 1951. The life of the factory began with the production of AZ1 using production equipment obtained as a reparation of the Second World War. Mass and serial production of tubes was based on Philips technology and production equipment under a license agreement with Philips in 1959. The installed capacity was 12,000,000 tubes per year. The product range was 126 types of tubes (Noval, Octal, Magnoval). During the years of operation of the factory, the production equipment has been well maintained and upgraded with original equipment developed and manufactured by the Development department of Ei Radio Tube Factory. Also, many technological innovations that have been applied in production processes of tubes have been developed, as well as many production tools during the long life and work of *this factory. A lot of tubes with the original logo and brand name of the most famous tube factories were produced in this factory after the discontinuation of their production, especially after 1970. One of the last tube developed in this factory was the famous KT 90 (developed by the tube designer Mr. Blaza Bukumira). The author of the book believes that the KT 90 is the best beam tetrode ever produced. After the political and economic transition in Serbia at the end of 20th century, factory equipment was sold to Western Electric (2006). The author of the book had the idea and hope that Western Electric would re-establish the production of tubes from the product range of the Ei Radio Tube Factory, with some modernization and upgrading of production equipment and improvement of technological processes. (It was one of the professional missions of the author of the book as vice president and deputy general manager of Electronic industry Nis. The author of the book wished the success of this mission as the crown of his professional career and due to the fact that he started his professional career in Ei Radio Tube Factory and participated in several reengineering of this factory). The author wanted to design an amplifier in gratitude to Western Electric for restarting the production of Ei tubes. Unfortunately, the good and noble intentions of the author and the good will of Mr. Charles Whitener, CEO of Western Electric did not have a happy ending. Much of the Ei Radio Tube factory equipment owned by Western Electric has been relocated or sold, some of it was damaged and devastated. Sad end of a Ei Radio Tube factory.*

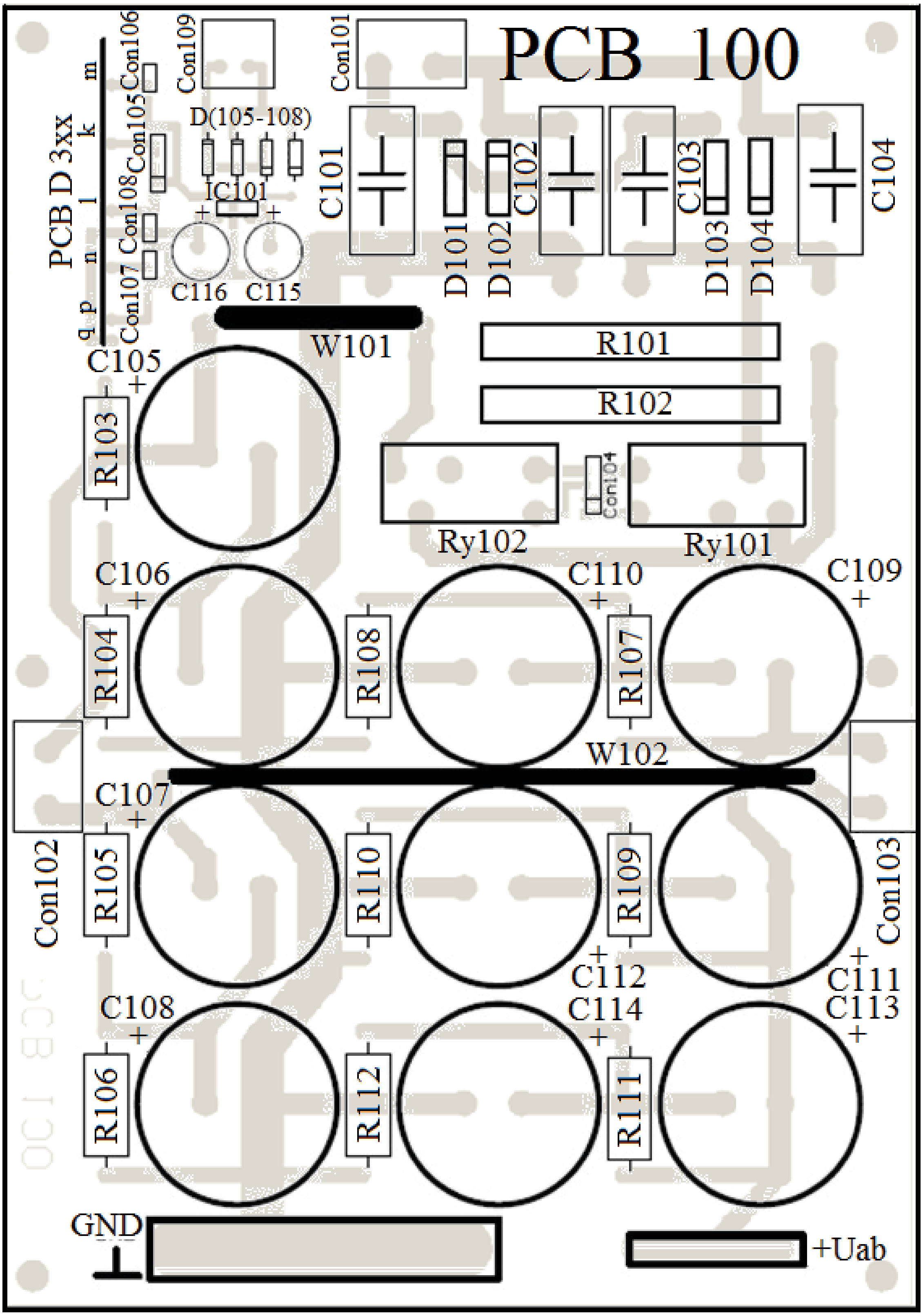
** The amplifier was made using WE437A and WE300B tubes given to the author by Mr. Charles Whitener, CEO of Western Electric. Thank You Charles.*
The left and right channel of the amplifiers are named after the author's granddaughters: Masha and Dunja, (the MM RIAA preamplifier with Ei EC900 is named after the third granddaughter Sena).

Either way the amplifier is designed and built and perhaps some design details may be useful to amplifier designers.

SSE 2 × 300B schematic

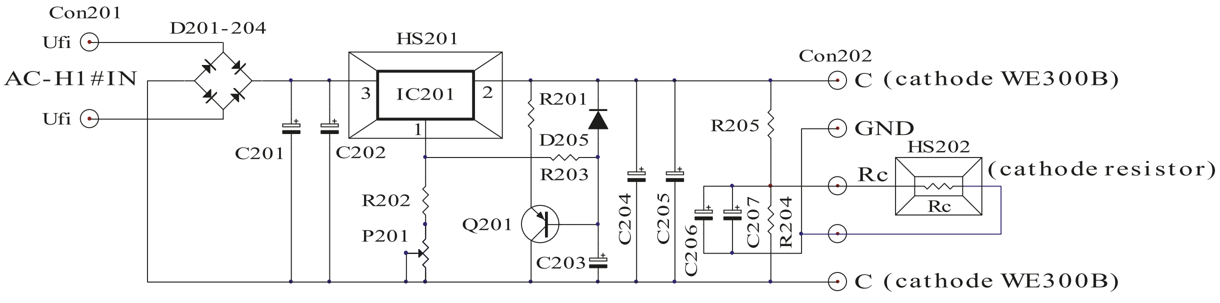


Power supply (main board)



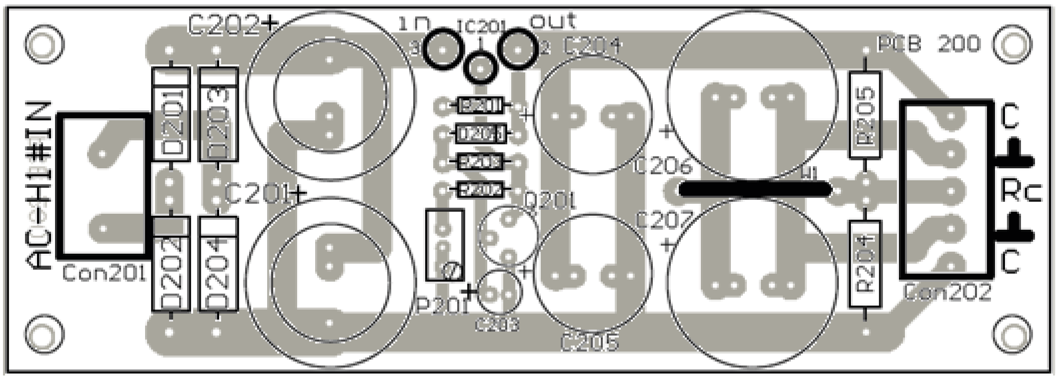
 # Driver and input stage circuits high voltage power supply: Regulated (Maida type) DC power supply.
 # WE 437A and KT88 heater power supply: Regulated DC power supply (LM 337).

 # Output tubes heater power supply (separately for each WE 300B tube): Regulated DC power supply (LM 338) It is combined with an output tube automatic bias circuit: cathode resistors and bypass capacitors.

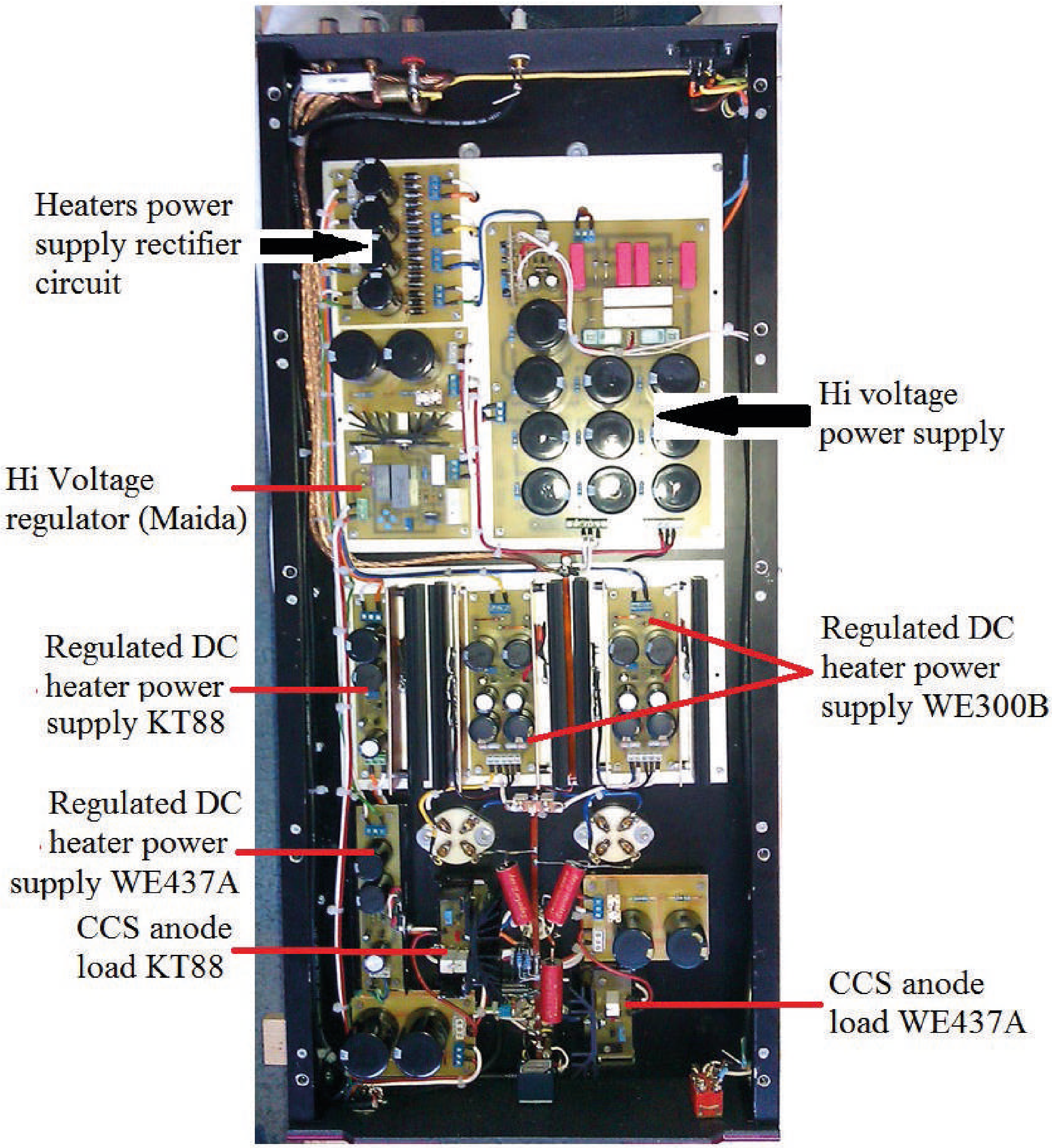


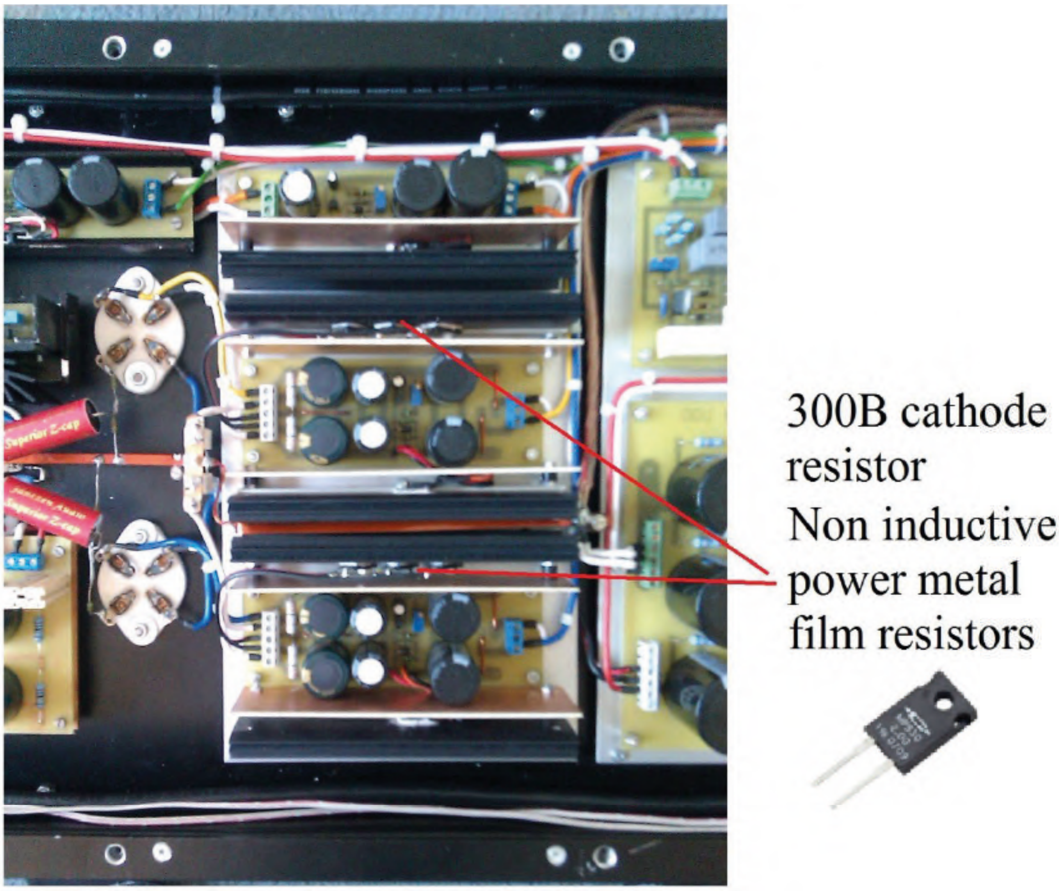
DC3 00B HEATING - PCB 200 / SE300B
PCB200

R201	120 Ω / 0.25W	D201	1N5402	C201	10.000 µF / 25V	Q201	BC557B
R202	120 Ω / 0.25W	D202	1N5402	C202	10.000 µF / 25V	IC201	LM338
R203	100 kΩ / 0.25W	D203	1N5402	C203	100 µF / 16V	P201	500 Ω
R204	22 Ω / 2W	D204	1N5402	C204	220 µF / 350V	Rc (200+200+500) Ω	
R205	22 Ω / 2W	D205	1N4002	C205	220 µF / 350V	MP 820	



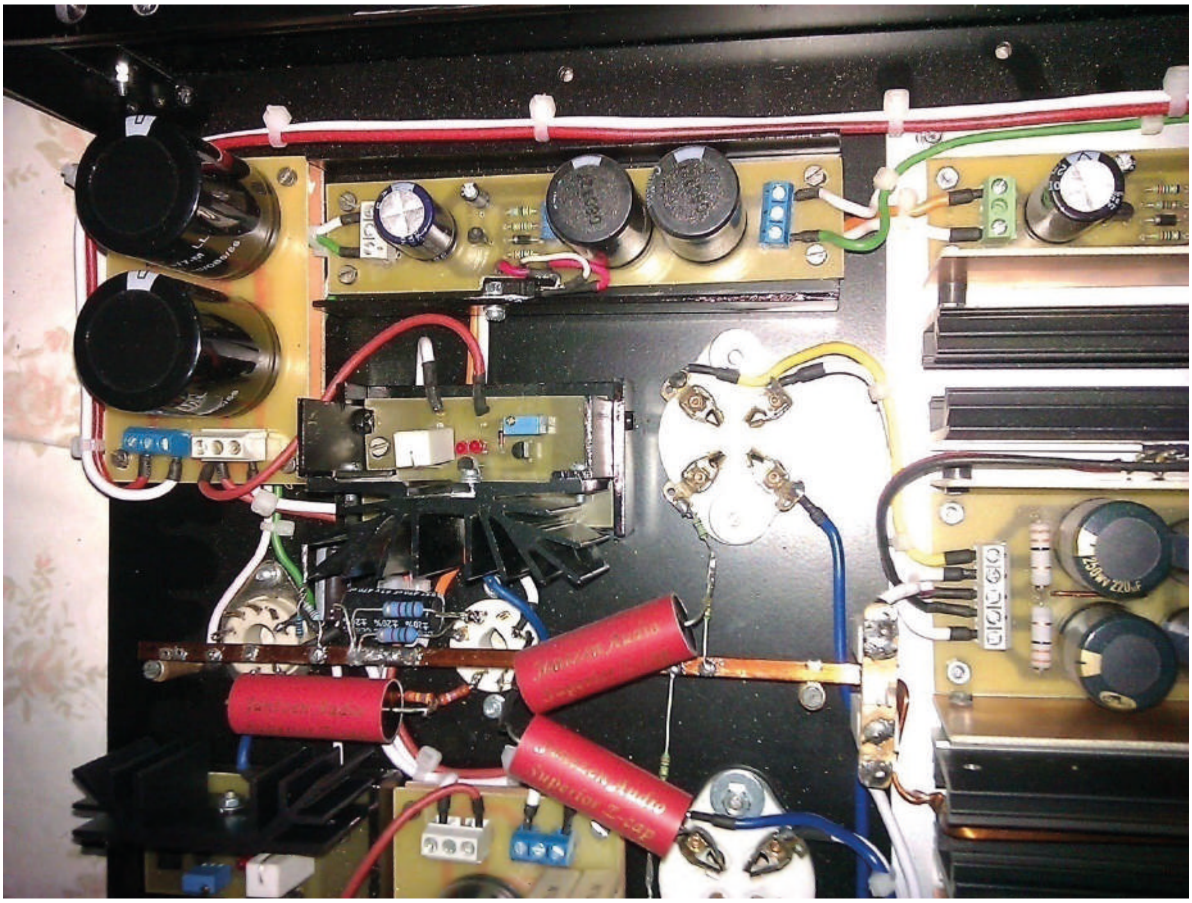
Dimension: (140 × 50) mm

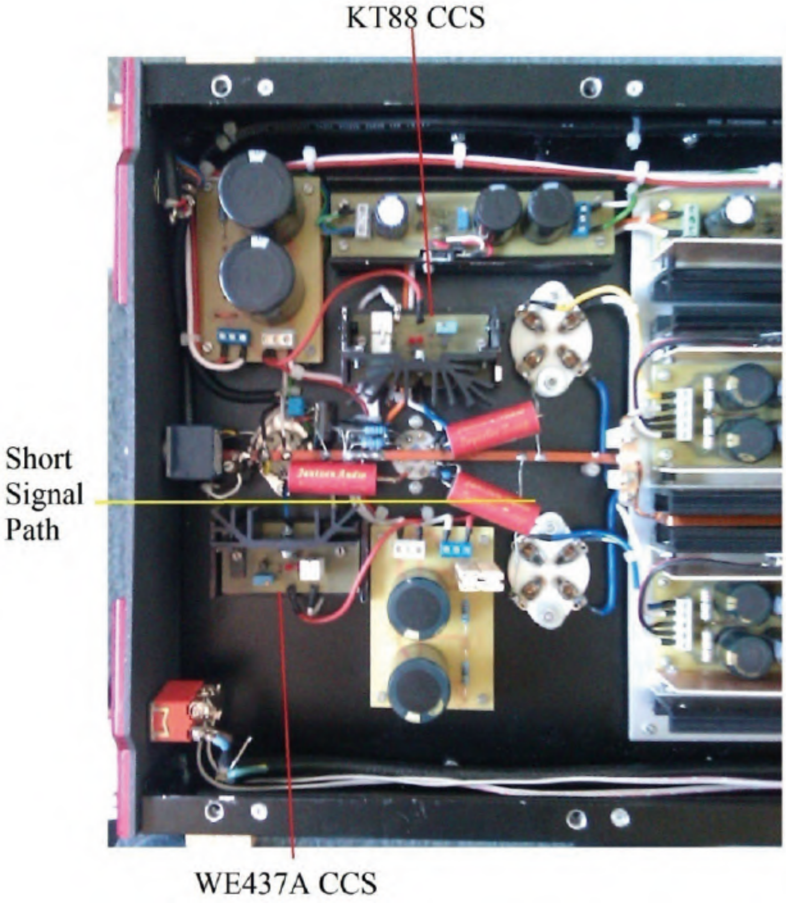

Heaters power supply rectifier circuit
Hi voltage power supply
Hi Voltage regulator (Maida)
Regulated DC heater power supply KT88
Regulated DC heater power supply WE300B
Regulated DC heater power supply WE437A
CCS anode load KT88
CCS anode load WE437A



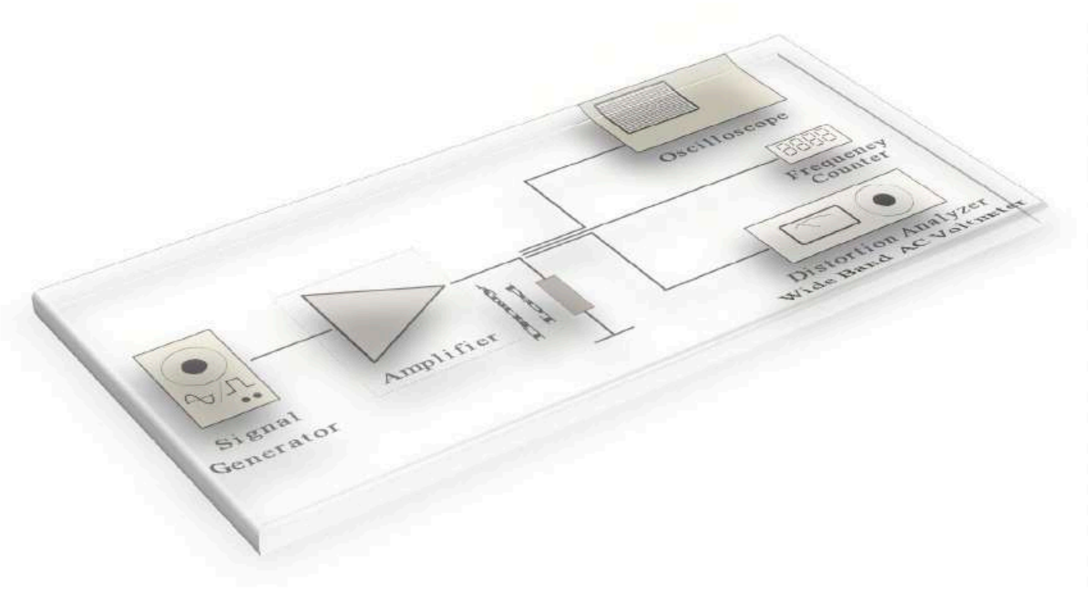
300B cathode resistor

Non inductive power metal film resistors




KT88 CCS
Short Signal Path
WE437A CCS

Chapter 8 • Measurement and Testing



Introduction Note

The measurement process begins long before the amplifiers building process. Each electronic, electromechanical and mechanical component must pass a so-called " incoming inspection ". This means that each component must be checked visually, dimensionally and in terms of characteristic published in the technical datasheet.

Visually - the component must not have visible mechanical damage, shape deformation or discoloration. Measuring instrument: human eye.

Dimensionally - Physical dimensions must be within tolerances published in a data sheet. Measuring instruments: Instruments for measuring length (calipers, micrometer, graduated ruler, measuring tape), weight (precision scale), ...

Electrical characteristics

- Passive electronic components: capacitors and resistors can be relatively easily measured using measuring instruments such as capacitance meter and resistance meter (in practice usually: universal multimeter (digital or analog – DC and AC current and voltage, Ohm meter, Capacitance meter). Capacitance and resistance can be measured by direct measuring method (specific instruments) or by indirect measuring method (instruments for measuring DC and AC current and voltage). In a case of the third passive component – inductivity, measurement is complicated. There are instruments for direct measurement of inductance (RLC meter), but they are effective only in the case of measuring the inductance of air coils and coils with ferrite cores that operate with low magnetic flux. In the case of measuring the inductance of power transformer coils or output transformers coils and chokes with transformer laminated core, and in general inductive components that operate with high magnetic flux, many methods of indirect measurement have been developed.
- The characteristics of active electronic components: vacuum tubes, transistors, FETs, can be measured by special equipment such as characteristic testers and curve tracer. Many indirect measurement methods can be used.
- Electro mechanic components: switches, relays, connectors..., - electromechanical functionality is tested.

It should be noted that it is necessary to carry out the procedure of checking each component after its installation in the amplifier due to the possibility of physical damage or change in the characteristics of the electronic component during the installation process (for example: physical damage of component or changing its electrical characteristics due to exposure to high temperatures during the soldering process).

The procedure of visual and ohmic checking of each solder junction is necessary.

All measurements should be carried out with great care and patience as accurately as possible. Measurement should be performed in room conditions (normal humidity $(30 \div 60)^0$ and temperature $(20 - 25)^0$C and in good lighting).

It is of the greatest importance that the measurement be objective (without the influence of our expectations, desires and emotions).

Basic measuring instruments:

3½ Digit Universal Multimeter

Typical Specifications:

- Voltage DC: 200 mV/2 V/20 V/200 V/1000 V ±(0.5%+3)
- Voltage AC: 200 mV /2 V/20 V/200 V/700V ±0.8%+5)
- Current DC: 2 mA/20 mA/200 mA/10 A ±(1.0%+3)
- Current AC: 2 mA/20 mA/200 mA/10 A ±(2.0%+3)
- Resistance: 200 Ω/2 kΩ/20 kΩ/200 kΩ/2 MΩ/20 MΩ/200 MΩ ±(0.8%+3)
- Capacitance: 20 nF/200 nF/2 μF/20 μF/200 μF ±(4.0%+3)
- Diode test: YES
- Triode test: YES
- On-off test: YES
- Auto power off: YES
- Data hold: YES
- Low Battery display: YES

Analog Universal Multimeter

Capacitance Meter

6 1/2 Digit Desktop Multimeter

RLC Meter

Testing and Measuring the Completed Amplifier

1. DC Measurement

1.a. DC Checking

Unloaded checking

- Tubes are **not inserted** in their sockets.
- Secondary coil of high-voltage of the power supply transformer is **not connected** to the rectifier circuit.
- All other secondary coils of the power supply transformer connected to their circuits (tube heaters, DC power supply circuit of the tube heaters, auxiliary circuit, **fixed bias circuit**,…).

Check the heater AC voltage at the tube socket pins.

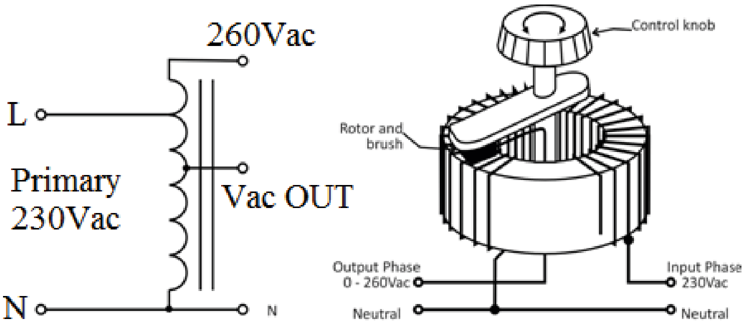
Check the heater DC voltage at the tube socket pins. Temporarily adjust the heater DC voltage to the nominal value (if the heater DC power supply circuit is adjustable). Temporarily adjust the DC voltage of the fixed bias circuit to the nominal value (if the fixed bias DC power supply circuit is adjustable).

Measuring instrument: Universal Multimeter.

 # High voltage DC power supply circuit - Checking (optional)

The high voltage DC power supply circuit can also be checked:

- *Secondary coil of high voltage of the power supply transformer is connected to the rectifier circuit.*
- *A power resistor (so – called **dummy load**) which simulates the ohmic resistance of the high DC voltage power supply line of the power amplifier circuit is connected to the output of the DC high voltage power supply circuit. The resistance and power of the resistor are calculated by applying Ohm's law.*
- *Example:*
  Amplifier high DC voltage power supply line: U = 430 V, I = 240 mA.
  $R_{Load\ DC}$ *= 430V / 0.24A = 1791Ω, standard value: 1k8. P =* U^2 */ R =* 430^2 */ 1800 = 102.7W, (it is necessary to use a resistor 5 times higher power ≈ 500W – 5 parallel connected wire wound resistor of 10 kΩ / 50 W).*



- *Amplifier connected to the mains via a **Variable Auto Transformer.***
  *An **auto transformer** is a type of electrical transformer with only one winding. This single winding is "tapped" at various points along its length to provide a percentage of the primary voltage supply across its secondary load.*



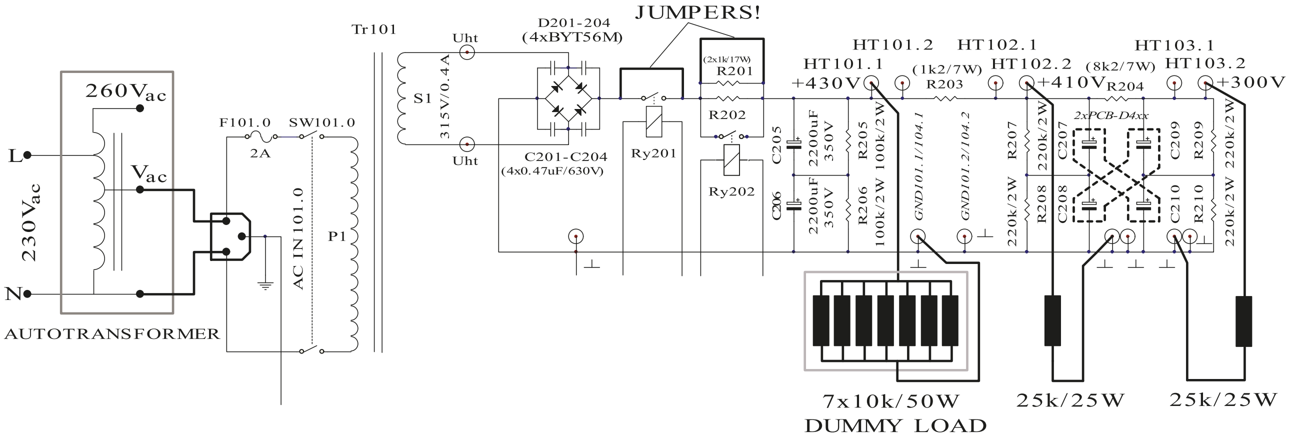


(It is most convenient to use an isolated variable AC Power Supply).

- Example: High voltage power supply of the PP amplifier.
  Procedure:
  - *Temporarily connect the wire jumpers across the time delay relays contacts.*
  - *Connect the amplifier to the autotransformer.*
  - *Slowly increase the voltage at the output of the auto transformer to the standard value of the mains supply voltage (230 Vac).*
  - *Continuously measure voltages at measuring points (HT101, HT102, HT103)*
  - *The measured voltage values at the measuring points should be close to the projected values at the end of the measuring i.e. the voltage at the output of the auto transformer = standard mains voltage.*

* The DC high voltage circuit is not connected to other amplifier circuits for the entire *duration of the measurement.*

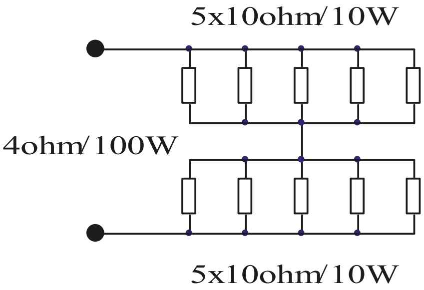
- *Switch off the amplifier and disconnect it from the auto transformer.*
- ***Remove (disconnect) wire jumpers from the delay circuit relays contacts.***
- ***Remove (disconnect) dummy loads.***
- ***Finish the amplifier wiring** (connect the DC high voltage circuit to the other amplifier circuits).*



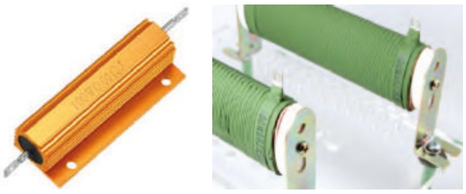
1.b. DC Measurement

DC measurement of the finished amplifier

- Place the tubes in the sockets.
- Set the volume potentiometers to min. position.
- Short - connect the input cinches using auxiliary tools:
- Connect the dummy load (4 Ω or 8 Ω) to the output of the amplifier:
  Dummy load can be made using non - inductive high power resistors



(2 Ω, 4 Ω, 8 Ω, … / 100 W), or using a combination of several lower power resistors:



- Turn on the amplifier and wait a few minutes while the amplifier warms up and the operating conditions stabilize.

- Measure the heaters AC voltages at the socket pins. **For maximum measurement accuracy, perform measurements with a multimeter set to the range closest to the expected voltage value.** ***Example: Push pull amplifier, output tubes KT88, heater supply 6.3 V_{AC}, socket pins 2 and 7. Multimeter: AC, range 20 Vac.***
  The measured voltage must be within the tolerances published in the tube data sheet.
  If the measured voltage is higher than the maximum allowable voltage, the problem can be fixed by inserting (in series) a high power resistor into the heater power supply circuit.

Example: KT88, Uf = 6.3 Vac, If = 1.6 Vac. Measured: Uf = 6.82 Vac.
Series resistor: Rs = ($Uf_{measured}$ − Uf) / If = (6.82 − 6.3) / 1.6 = 0.325 Ω, standard: 0.33 Ω. Resistor dissipation power: P = Rs × I^2 = 0.33 × 1.6^2 = 0.84 W.
Use a resistor few times higher power than the calculated above (5W). Rs: 0.33 Ω / 5 W.

Be sure to turn off the amplifier before performing this heater voltage correction procedure.

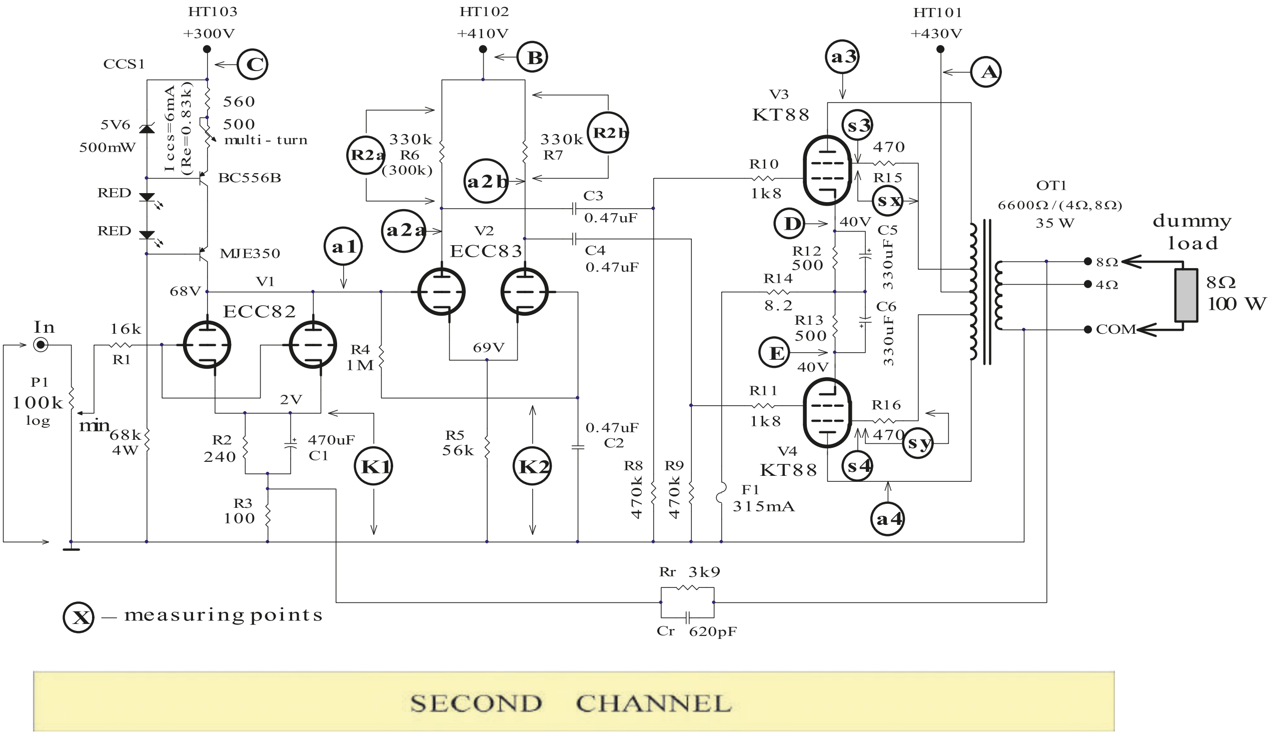
- Measure the heaters DC voltages at the socket pins.
  The procedure is the same as in the previous measurement, except that the multimeter is set to measure the DC voltage (20 Vdc range).
  The problem of higher voltage than tolerated can be solved as in the previous case (by inserting a resistor into the heater supply circuit).

  Be sure to turn off the amplifier before performing this heater voltage correction procedure.
  If an adjustable heater supply circuit is used, set the heater voltages to the nominal value.

- Measurement of the DC voltages along the high DC voltage power supply line:
  high DC voltage power supply of output stage, high DC voltage power supply of driver stage, high DC voltage power supply of input stage. *(In the above example of a push pull amplifier, measure the DC voltage at the measuring points:*

  HT101 (A), HT102 (B), HT103 (C)).

  Use a multimeter to measure DC voltage and set it to the appropriate measuring range.



Continue the measurement procedure by measuring the DC voltage at important measuring points of the circuit:

- Measuring points: **D** and **E** – Automatic bias voltage.
  The voltage at point D (E), in addition to the **quiescent point** information (U_D = $U_{g\text{-}k}$), also contains information about the output tube current: (I = $U_{measured}$ / $R_{cathode}$).

  By calculating the currents of the output tubes and comparing them, the unbalance ($U_D \neq U_E$) of the branches of the push pull circuit can be estimated. If a balancing circuit is used, make the necessary adjustments to achieve good balancing of the branches of the push pull output stage circuit.

Fixed bias voltage: Measure the voltage across the cathode test resistor. Apply the procedure of bias voltage adjustment and push pull output circuit branch balancing (use procedure of current calculating:

$I = U_{measured} / R_{test}$).

- Measuring points: **a3** and **a4**.
  By measuring the voltage at the anodes of the output tubes (measuring points: a3 and a4), the voltage between the anode and the cathode of each tube can be calculated:

  $U_{(a-k)3} = U_{a3} - U_D$, ($U_{(a-k)4} = U_{a4} - U_E$).

- Measuring points **s3** and **s4** provide information about screen grid voltage.
- By measuring voltage across the screen resistors (**sx**) and (**sy**), the screen current (I_{g2}) of each output tubes can be calculated*:* $I_{g2} = U_{(sx)} / R_{g2}$.
- ** After these measurements and adjustments, almost all parameters of the output tubes DC operation are known:* U_{a-k}, U_{g-k}, $I_a = I_k - I_{g2}$. *These parameters can be compared with tube manufacturer data sheet figures and diagrams.*
- Measuring points: **a2a**, **a2b** and **K2**
  By measuring the voltage at the anodes (a2a and a2b) and the voltage at the cathode (K2) of the phase splitter tube, the voltage between the anode and the cathode of each half of the tube can be calculated. By measuring the voltage across the anode resistor (**R2a**, **R2b**), the DC current of each half of the phase splitter tube can be calculated.

- Measuring points: **a1** and **K1**
  By measuring voltage at point K1 (voltage across the input tube cathode resistor = Ug-k), the anode current can be calculated and can be adjusted to the desired value by adjusting the CCS current.
  After measuring the voltage at the anode of the input tube – point a1, all the parameters that define the operating conditions of the input tubes and the phase splitter tube are known:
  Input tube: $U_{a-k} = U_{a1} - U_{K1}$, $I_a = U_{K1} / R_{k1}$, $U_{g-k} = U_{K1}$
  Phase splitter tube: $U_{g-k} = U_{a1} - U_{K2}$,
  Half of the tube a): $U_{a-k} = U_{a2a} - U_{K2}$, half of the tube b): $U_{a-k} = U_{a2b} - U_K$.
  Half of the tube a): $I_a = U_{R2a} / R_{a(a)}$, half of the tube b): $I_a = U_{R2b} / R_{a(b)}$.

The rules apply to all these measurements:

- Carry out the measurements with great care.
- Take care of safety (Caution: Voltages in the amplifier can be deadly).
- Beware of possible burns (some components can be very hot – over 60^0C).
- Use the optimal measuring range of the multimeter.

Repeat all measurements after 30 and 120 minutes of amplifier operation.

- 1.c. Temperature measurement

  Measuring the temperature of the power electronics components, measuring the temperature inside the chassis and measuring the temperature on the outer surface of the chassis under the operating condition of the amplifier is necessary.

  These measurements provide some necessary information:

  - Safety of amplifier operation.
  - Estimation of the defined rated power of the electronic components used.
  - Operating temperature inside the chassis. Estimation of the permissible temperature inside the chassis and influence on the operation of other electronic components of the amplifier.
  - Safe operation of the users (permissible temperature of the outer surface of the chassis – the surface with which the user comes into contact during the use of amplifiers, except tubes).

Temperature measurements can be performed using **contact** and **non-contact** thermometers.

In practice, simple thermometers based on thermocouples are most often used. Almost all modern multimeters have the option of measuring the temperature using a thermocouple sensor probe. A non-contact infrared thermometer can also be used.

Temperature measurements should be performed when the thermal processes are stabilized and when the amplifier reaches the operating temperature (at least 30 to 120 minutes from the moment the amplifier is turned on).

Temperature measurement with thermocouple probe
The thermocouple probe has to be in physical contact with the object of measurement.

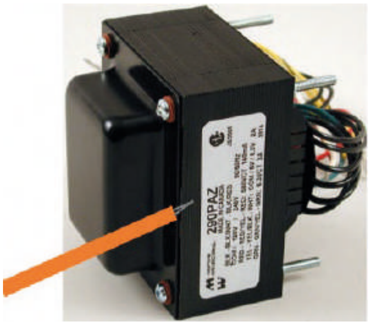
Example 1:
Power resistor temperature measurement.

The tip of the thermocouple probe must be placed on the surface of the resistor (it is good to apply thermal conductivity paste on the surface of the resistor at a measuring point to ensure good thermal conductivity at the junction of the thermocouple probe and the measuring object). The rated power of the resistor is specified by the manufacturer usually at an operating temperature of 70 ^{0}C. So, it is good that the operating temperature of the resistor is around 70 ^{0}C (preferably below 60 ^{0}C). If the measured temperature is well above 70 ^{0}C , it is a sign that the value of the resistor in terms of rated power is not well determined - in this case a higher power resistor must be chosen or the resistor must be replaced by a combination of two or more resistors (in series or parallel connection) with higher total power.

Temperature measurements should be performed:
resistors in hi voltage power supply line, cathode resistors of output tubes in case of automatic bias, resistors in the power supply circuit of the tube heaters, voltage distributors circuit and sometimes resistors in the tube anodes circuit, ...

Example 2:
Power transformer temperature measurement.

- *- Iron core temperature measurement*



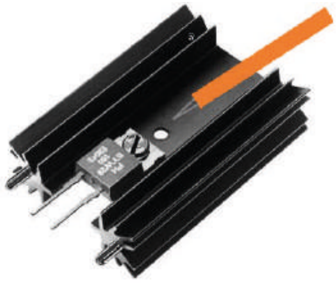
(In practice, it often happens that the surface temperature of the transformer in operation at rated power declared by the transformer manufacturer is 70 ^{0}C to 85 ^{0}C.
The most common reason is: the engineering calculation of the transformer is based on the transformer overheating of 60 ^{0}C and ambient temperature of 25 ^{0}C.The reasons are economic not technical).

Criteria for designing transformers used in audio amplifiers are much stricter.

Place the tip of the thermocouple probe on the iron core surface (it is good to apply thermal conductivity paste on the surface of the transformer) and read the operating temperature of the transformer on the thermometer (or multimeter set to measure the temperature). As transformers are most often mounted on the outer surface of the chassis, it is desirable that the temperature of the iron core does not exceed 60 ^{0}C.

(The most common reason of transformer overheating is caused by economic limitations – the minimum amount of material used to manufacture transformers, especially the amount of Cu wire. For example, if the primary current of an unloaded transformer is small and the measured temperature on the core surface of the loaded transformer (in operation) is high, it can be concluded that the transformer overheating is caused by overheating of transformer windings — insufficient cross-section wire gauge was used and heat energy got transferred to the transformer core).

Example 3:
Heat sink temperature measurement.



Place the thermocouple probe on the heat sink surface near the component mounted on the heat sink and read the operating temperature on the thermometer.

The maximum allowable temperature of the heat sink depends on several factors: the thermal characteristics of the component mounted on the heat sink and the allowed temperature in the environment of the heat sink, mainly.
The measured values can provide useful information on whether the heat sink was chosen correctly.

Again, it is good that this temperature is below 60 ^{0}C.

Example 4:
Chassis outer surface temperature measurement.

Measure the temperature at several points on the surface of the chassis, even near the tubes. Also measure the temperatures of the components mounted on the chassis (except tubes).
It is good that the temperature of the chassis parts that come in contact with the user during normal use of the amplifier is below 45 ^{0}C.

Example 5:
Measuring the operating temperature inside the chassis.
Measure the temperature at several points inside the chassis - that is actually the ambient temperature inside the chassis.
As many electronic components are sensitive to high temperatures (electrolytic capacitors, for example), it is good that the temperature inside the chassis is below 60 ^{0}C.

2. AC Measurement

Measurement of amplifier characteristics such as:

- Rated output power
- Harmonic distortion
- Frequency response
- Signal-to-noise ratio
- Input sensitivity
- Gain

provides information on the quality of the amplifier.

Measurement of the above characteristics of amplifiers is technically demanding and requires the use of specific measuring equipment:

- Audio oscillator (sine / square wave signal generator)
- Oscilloscope
- Distortion meter
- Frequency wideband AC voltmeter

  - Dummy load
  - Specific filters networks
  - Spectrum analyzer
  - Quality connecting cables

Tables, measurement sheets, diagrams, etc. are used to monitor the measurement process as well as for numerical and graphical presentation of measurement results.

The quality of the measuring process and the quality of the measurement results largely depend on the quality of the measuring equipment:

- Signal generator
  The signal generator should provide a low harmonic distortions sinusoidal signal in the frequency range covering the audible range (20 Hz – 20 kHz), and sufficient amplitude of the output signal necessary for full excitation of the measured audio amplifier. For more advanced measurements, the signal generator should be able to generate a square wave signal. Satisfactory measurement results can be achieved by using a signal generator with the following technical characteristics:

  - Sinusoidal and square wave signal
  - Frequency range: 10 Hz – 1 MHz
  - Output signal up to 10 V_{RMS}.
  - Output impedance: 50 Ω and 600 Ω.
  - Total harmonic distortion (THD): 0.02% or lower.

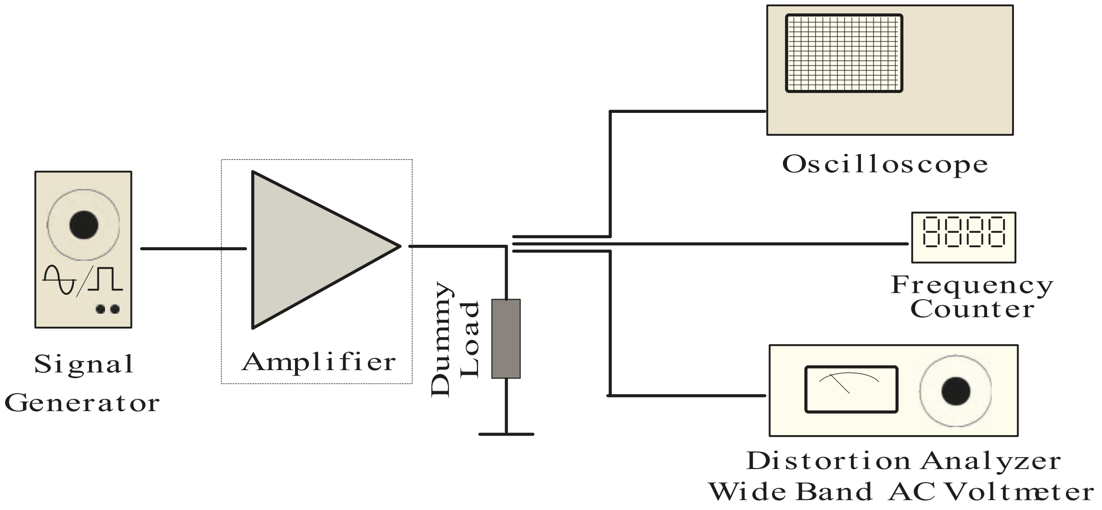
- Oscilloscope
  - Dual trace 10 MHz.

- Distortion Analyzer
  - Voltage measurement range 300 μV_{RMS} to 300 V_{RMS} (Full range).
  - Residual noise below 30 μV_{RMS}.
  - Voltage RMS and dB scale.
  - 10 Hz to 500 kHz range.
  - Measures Total Harmonic Distortion: 0.01% or lower.

- Accessories
  - Dummy load: (4 Ω and 8 Ω) / 100 W.
  - Quality audio cables and measuring probes.
  - Short – circuit input connector.
  - Frequency counter
  - Specific filter networks (Inverse RIAA, for example).

Test Setup
Notes:
Measurements at home, unlike laboratory measurements, do not take place in controlled conditions inside a Faraday cage that blocks external electromagnetic fields. Home appliances, elevators in buildings, work activities in the environment, welding, dirty mains supply, ... can be the cause of unwanted radiation and interference that can greatly compromise the quality of measurements at home. To reduce the impact of these disturbances, measurements can be made on the part of the day when these effects are smallest. ***Also due to safety measures but also due to the quality of measurements, all instruments used must be grounded in an appropriate manner.***

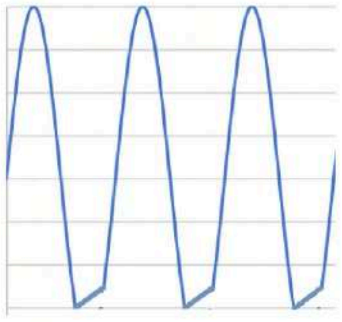
Measurement system configuration
- Connect the output of the signal generator to the input of the amplifier using a quality cable.
- Connect the dummy load of appropriate resistance (4 Ω, 8 Ω, ...) to the amplifier output.
- Connect the oscilloscope probe (one channel) to the amplifier output (dummy load).
- Connect the oscilloscope probe (other channel) to the amplifier input.
- Connect the distortion analyzer probe to the amplifier output.



Output power test procedure

- Signal generator set to a sine wave of 1 kHz and a minimum output level.
- Amplifier volume control set to maximum.
- Turn on the amplifier and all measuring instruments.
- Wait 30 minutes necessary to achieve optimal operating conditions of the amplifier and measuring instruments.
- Carefully increase the level of the output signal of the signal generator and at the same time visually monitor the image of the output signal of the amplifier on the oscilloscope screen.
- Depending on the level of the output signal, change the measuring ranges of the oscilloscope and AC voltmeter of the distortion analyzer in a timely manner.
- Increase the output signal of the signal generator until it is visible on the oscilloscope screen that the sine wave signal has started to clip:

Clipping at the top and bottom



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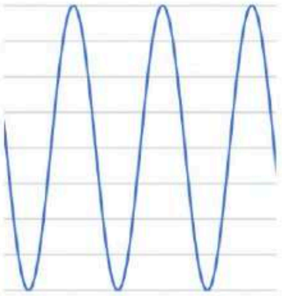


Maximum level without clipping

- Slightly reduce the signal at the input of the amplifier until the sine wave displayed on the oscilloscope screen returns to the shape of an undistorted sine wave.
- Read the RMS voltage at the output of the amplifier on the AC voltmeter of the distortion analyzer.
- Calculate the output power of the amplifier:

$$P_{out} = \frac{U_{RMS}^2}{R_{Load}}$$

- The power measured and calculated in this way can be accepted (conditionally*) as the **maximum output power** of the amplifier.

*) A difference should be made between the maximum output power and the rated output power of the amplifier. THD at rated power must be lower (or equal) than the rated maximum THD at all frequencies within the rated power bandwidth.

Within this measurement, other characteristics of the amplifier can be determined:
amplification and maximum input signal **(sensitivity** for P_{max}**)** of the amplifier.

- Simultaneously with the measurement of the output voltage of the amplifier ($U_{out-RMS}$), read the output level of the signal from the signal generator or measure the RMS voltage at the input of the amplifier using an AC voltmeter (U_{in-RMS}).

- Calculate the voltage amplification of the amplifier:

$$A = \frac{U_{out-RMS}}{U_{in-RMS}} \quad \text{or express it } [dB]: \; A[dB] = 20 \times \log \frac{U_{out-RMS}}{U_{in-RMS}}$$

These measurements determined the following set of the amplifier characteristics:

- **Maximum power: P_{max} = _ _ _ [W], (1 kHz, undistorted signal)**

- **Sensitivity for P_{max} : $U_{in\,(Pmax)}$ = _ _ _ [V_{RMS}]**

- **Amplification: A = _ _ _ [x], A = _ _ _ [dB]**

Frequency Response or Amplitude Characteristic
The amplitude characteristic of the amplifier is the amplification as a function of the frequency of the input signal:

$$A = \frac{U_{out}}{U_{in}} = f(frequency)\,\Big|_{U_{in=ctc}} \quad or, \qquad A(dB) = 20 \times log\frac{U_{out}}{U_{in}} = f(frequency)\,\Big|_{U_{in=ctc}}$$

- The configuration of the measuring system is the same as in the previous measurement.
- Measurement can start under the same conditions as in the previous measurement i.e. maximum amplifier power and 1kHz input signal.
- One of the results of the previous measurement is the amplification of the amplifier (1 kHz input signal) The amplification is expressed as the ratio of output and input RMS voltage ($U_{out-RMS}$ / U_{in-RMS}) or (20 × log ($U_{out-RMS}$ / U_{in-RMS}) [dB]) at an input signal of **1 kHz.** The measured voltage ($U_{out-RMS}$) at the output of the amplifier under the above conditions represents the **reference level (U_{REF})** in the further measurement procedure.

Measurement procedure:

- Read the RMS voltage at the output of the amplifier on the AC voltmeter of the distortion analyzer (or read it on dB scale).
- Slowly decrease the **frequency** of the output signal of the signal generator (frequency of the input signal of the amplifier) without changing its level (**keep the level of the input signal constant**) up to 500 Hz.
- Read the RMS voltage (at 500 Hz) at the output of the amplifier on the AC voltmeter (or dB) and enter this data in the previously prepared table.
- Repeat the previous procedure using signal frequencies of: 250 Hz, 100 Hz, 80 Hz, 60Hz, 40 Hz and 20 Hz, respectively, up to the lowest possible frequency the signal generator can produce (usually 10 Hz).
- Reset the signal generator to a frequency of 1 kHz.
- Slowly decrease the frequency of the output signal of the signal generator and at the same time visually monitor the voltage at the output of the amplifier on the AC voltmeter until the RMS voltage drops to 0.707 × $U_{out-RMS\,(1kHz)}$ or −3 dB from the reference level.
  Read the signal frequency of the signal generator – this is the **cut-off low frequency** (**f_L**).
  **These measurements can be performed if the cut-off low frequency is higher than the lowest frequency of the signal generated by the signal generator used.*
- Repeat the above set of measurements but using signal frequencies of 2 kHz, 5 kHz, 10 kHz, 12 kHz, 16 kHz, 20 kHz, 30 kHz, 40 kHz, 50 kHz, 60 kHz, 70 kHz, 80 kHz, 90 kHz, and 100 kHz.
  Cut-off high frequency (**f_H**): perform the same procedure as in the previous measurement set.

**** Keep the signal at the input of the amplifier at a <u>constant level</u> throughout the measurement procedure.***

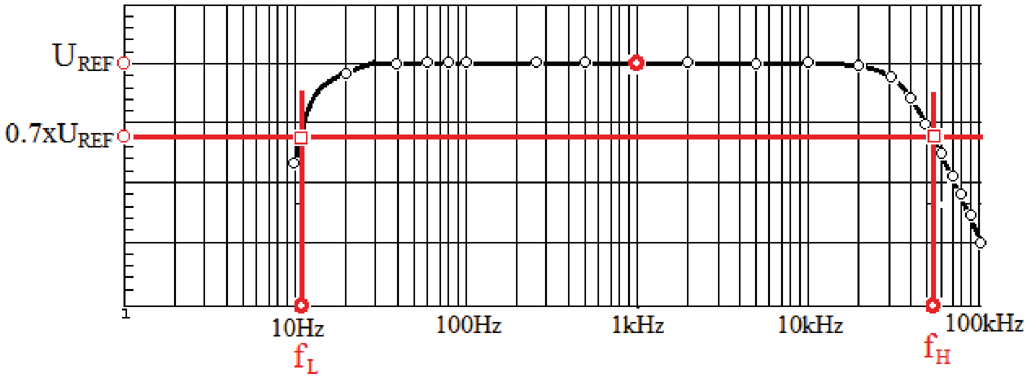
Uin = - - -	10Hz	**20Hz**	40Hz	60Hz	80Hz	100Hz	250Hz	500Hz	**1kHz**
Uout									
dB									
Uout / Uin									
20log(uout/Uin)									

- Cut-off low frequency f_L = _ _ _

Uin = - - -	**1 kHz**	2 kHz	5 kHz	10 kHz	12 kHz	16 kHz	**20 kHz**	30 kHz	40 kHz	50 kHz	60 kHz	70 kHz	80 kHz	90 kHz	100 kHz
Uout															
db															
Uout / Uin															
20log(uout/ Uin)															

- Cut-off high frequency f_H = _ _ _

Graphic presentation of numerical measurement results



The complete previous procedure of measuring the frequency response of the amplifier should be repeated at the following levels of the output signal of the amplifier:

- $U_{out-RMS} = 0.9 \times U_{out\,(max)\,-RMS}$ or – 0.9 dB, ($P_{out} = 0.8 \times P_{out\,(max)}$).
- $U_{out-RMS} = 0.7 \times U_{out\,(max)\,-RMS}$ or – 3 dB, ($P_{out} = 0.5 \times P_{out\,(max)}$).
- $U_{out-RMS} = 0.5 \times U_{out\,(max)\,-RMS}$ or – 6 dB, ($P_{out} = 0.25 \times P_{out\,(max)}$).
- P_{out} = 1W (R_L = 4 Ω, U_{out} = 2 V_{RMS}; R_L = 8 Ω, U_{out} = 2.83V_{RMS}).
- P_{out} = 250mW (R_L = 4 Ω, U_{out} = 1 V_{RMS}; R_L = 8 Ω, U_{out} = 1.41 V_{RMS}).
- ***$P_{out\,(rated)}$, rated power** of the amplifier: the **specified output power of the amplifier at a certain level of distortion (THD).**

** Another set of output levels (amplifier output power) can be selected, as well as a set of frequencies to be used to measure the frequency response, depending on the desired measurement accuracy or some other reasons.*

Example:
Power Amplifier: $P_{out\,(max)}$ = 14 W

Frequency response measurement sets:

- *Characteristic 1): P_{out} = 12 W.*
- *Characteristic 2): P_{out} = 10 W.*
- *Characteristic 3): P_{out} = 7 W.*
- *Characteristic 4): P_{out} = 5 W.*
- *Characteristic 5): P_{out} = 3 W.*

- *Characteristic 6): P_{out} = 1 W.*
- *Characteristic 6): P_{out} = 0.25 W.*

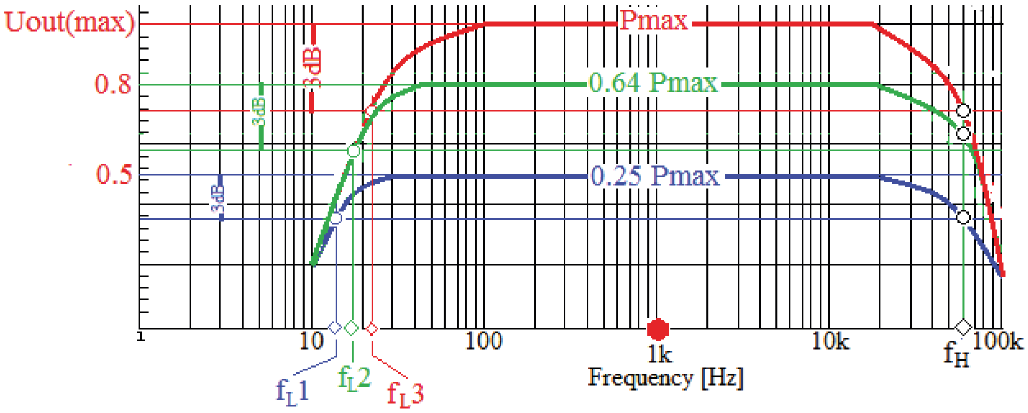
at a signal frequencies: 10 Hz, 20 Hz, 40 Hz, 60 Hz, 80 Hz, 100 Hz, 250 Hz, 400 Hz, 1 kHz, 2 kHz, 3 kHz, 5 kHz. 8 kHz, 10 kHz, 12 kHz, 16 kHz, 20 kHz, 25 kHz, 30 kHz, 40 kHz, 50 kHz, 60 kHz, 70 kHz, 80 kHz.

Example:
Power Amplifier: $P_{out\ (max)}$ = 35 W

Frequency response measurement sets:

- *Characteristic 1): P_{out} = 30 W.*
- *Characteristic 2): P_{out} = 20 W.*
- *Characteristic 3): P_{out} = 15 W.*
- *Characteristic 4): P_{out} = 5 W.*
- *Characteristic 6): P_{out} = 1 W.*
- *Characteristic 6): P_{out} = 0.25 W.*

at signal frequencies: 10 Hz, 20 Hz, 100 Hz, 500 Hz, 1 kHz, 10 kHz, 16 kHz, 20 kHz, 30 kHz, 40 kHz, 50 kHz, 60 kHz, 70 kHz, 80 kHz.
All sets of frequency response measurements (characteristics) can be presented and drawn in one common graph:



The above measurements determined the following set of amplifier characteristics:

- **Frequency response: (f_L = _ _ _, f_H = _ _ _ Hz), at P_{max} = _ _ _ W**
- **Frequency response: (f_L = _ _ _, f_H = _ _ _ kHz), at P_{rated} = _ _ _ W**
- **Frequency response: (f_L = _ _ _, f_H = _ _ _), at P_{1W} = 1 W**
- **Frequency response at P_{rated} = _ _ _ W :**
  (f = 20Hz, - _ _ _ dB), (f = 20kHz, - _ _ _ dB).

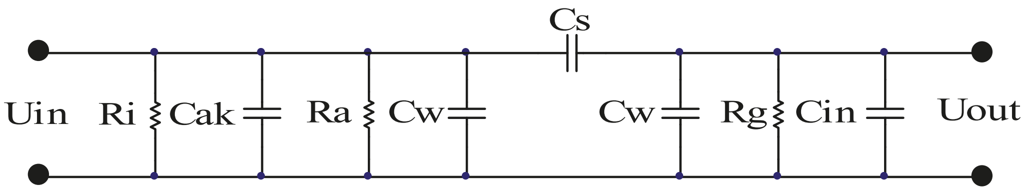
** These measurements not only serve to specify the frequency response of the amplifier, but also provide useful information to the amplifier designer on the quality of the amplifier design.*

Example:
By analyzing the results of measuring the frequency response of the amplifier, an unacceptably high cut-off low frequency was observed and a red alert was activated in the head of the amplifier designer. The question is: what is wrong with the design of the amplifier, what are the causes of poor frequency response of the amplifier?

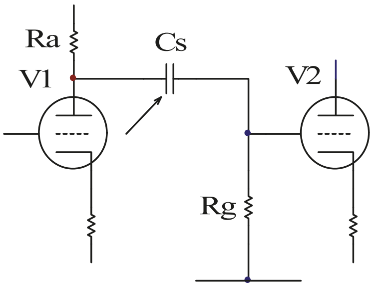
List of possible causes that affect the frequency response of the amplifier, especially at low frequencies:

- *If amplifier stages with RC coupling are used in the design of the amplifier, the possible cause of the high cut-off low frequency is the **low (i.e., insufficient) capacitance of the coupling capacitor**.*

  Reminder:
  *Amplifier with RC Coupling: the frequency **band-pass** filter exists between two stages.*



Low frequencies:



Cut-off low frequency (f_L):

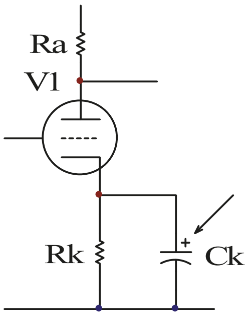
$$f_L = \frac{1}{2 \times \pi \times \boldsymbol{C_S} \times \left(R_g + \frac{R_i \times R_a}{R_i + R_a}\right)}$$

** If the amplifier has several RC couplings, the net cut-off low frequency can be calculated using a practical approximate equation:*

$$f_{L-net} = \sqrt{(f_{L1})^2 + (f_{L2})^2 + \cdots + (f_{Ln})^2}$$

where f_{Li} is f_L of each amplifier stage.

$f_{L\text{-}net}$ is higher than the highest cut-off low frequency of each amplifier stage.



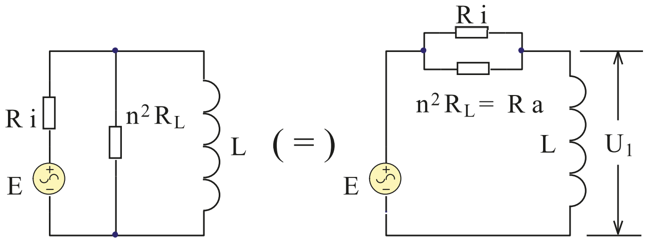
- If Grounded Cathode Stages with Bypassed Cathode Resistor are used in *the design of the amplifier, the possible cause of the high cut-off low frequency is the* ***low (insufficient) capacitance of the cathode resistor bypass capacitor****.*

  Cut-off low frequency:

$$f_{L(-3dB)} = \frac{1}{2\times\pi\times R_{keq}\times \boldsymbol{C_k}} \qquad R_{k\,eq} = R'_k \| R_k = \frac{\frac{R_i+R_a}{\mu+1}\times R_k}{\frac{R_i+R_a}{\mu+1}+R_k} \qquad R'_k = \frac{R_i+R_a}{\mu+1}$$

- *The possible cause of the high cut-off low frequency is the poor design of the amplifier* ***Output Transformer:***
  low or insufficient ***inductivity of the primary coil****.*

  Reminder:
  Output transformer equivalent electric diagram at low frequencies



$$f_{Low(-3dB)} = \frac{R_i \times R_a}{R_i + R_a} \times \frac{1}{2 \times \pi \times \boldsymbol{L}}$$

Inductivity of the Primary coil:

$$L_P = \frac{1.256 \times \boldsymbol{S_{eff}} \times \mu_e \times \boldsymbol{N_P^2}}{10^8 \times l_{Fe}}$$

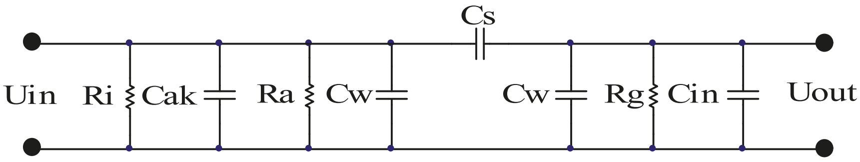
The cause of insufficient inductance of the Primary coil (L_P) can be:

- *insufficient number of turns of the Primary (inductance is proportional to the square of the number of turns of the Primary).*
- *insufficient cross section (S_{eff}) of the transformer core (inductance is directly proportional to the cross section of the core).*

Example:
By analyzing the results of measuring the frequency response of the amplifier, an unacceptably low cut-off high frequency (f_H) was observed.

List of possible causes that affect the frequency response of amplifiers, especially at high frequencies:
In general, the cut-off high frequency of the amplifier is determined by the capacitance or inductance associated with each electronic component, the layout of the electronic components and the electrical wiring inside the amplifier.

Audio amplifiers, in most cases, have a multi-stage configuration with vacuum tubes as active components.
The output impedance of the previous stage and the input impedance of the next stage form a band-pass filter:



$$f_H = \frac{1}{2 \times \pi \times C_{eq} \times R_a'}$$

$$\frac{1}{R_a'} = \frac{1}{R_i} + \frac{1}{R_a} + \frac{1}{R_g};$$

$C_{eq} = C_{ak} + C_w + C_w + C_{in}$, *C_W — (stray) capacitance caused by component wiring.

$C_{in} = C_{gk} + C_{ak} \times (A + 1)$; *Miller effect*

* *If the amplifier has several RC couplings, the net cut-off high frequency can be calculated by using a practical approximate equation:*

$$f_{H-net} = \frac{1}{\sqrt{\frac{1}{(f_{H1})^2} + \frac{1}{(f_{H2})^2} + \ldots + \frac{1}{(f_{Hn})^2}}}$$

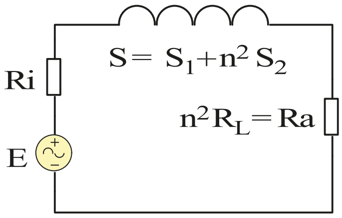
where f_{Hi} is f_H of each amplifier stage.

$f_{H\text{-}net}$ is lower than the lowest cut-off high frequency of each amplifier stage.

The frequency response of amplifiers at high frequencies depends on the characteristics of the vacuum tubes used (R_i , μ, C_{gk} , C_{ak}), the operating conditions of the tubes (A), the number of amplifier stages, the layout of the electronic components and electrical wiring inside the amplifier.

*One of the reasons for the poor frequency response of amplifiers at high frequencies may be the inadequate construction of the **Output Transformer**.*

Output transformer equivalent electric diagram at high frequencies:



$$f_{High(-3dB)} = \frac{R_i + R_a}{2 \times \pi \times S}$$

S: Leakage inductance

** The characteristics of the output transformer at high frequencies mostly depend on the geometry of the coils and number of turns of the coils (leakage inductance, stray capacitances) and magnetic characteristics of the core laminate material and laminate thickness.*

Total Harmonic Distortion (THD) or Total harmonic Distortion + Noise - (THD+N)

Reminder

Generally, distortion can be defined as a measure of the deviation of the shape of the signal at the output of the amplifier compared to the shape of the signal at the input of the amplifier.

- ***Non linear distortion***
  - ***Harmonic*** *(if the input signal is a periodic function in time, the output signal of the amplifier contains a fundamental harmonic and harmonics of higher order).*
  - ***Non harmonic (****nonlinear non harmonic distortions occur when the amplifier is driven by more than one AC signal. Anode current contains complex signal combined by the AC signals at the input of the amplifier: fundamental harmonic and harmonics of higher order, sum and subtractions of the fundamental harmonic and harmonics of higher order caused by the nonlinear characteristic of the tube).*

- ***Total Harmonic Distortion – THD***

  THD is a measure of the total value of all harmonics produced at the output of the amplifier – other factors are excluded. The amplitudes of the harmonics at the output of the amplifier are measured, their RMS value is added and the relation to the output signal is formed. The value of the Total Harmonic Distortion (THD) is the ratio of the square root of the sum of the squares of the values of the harmonics (RMS) at the output of the amplifier and the square of the RMS level of a sinusoidal output signal.

The THD is usually expressed as distortion in percentage ***(%)*** *or in dB - relative to the level of fundamental harmonic as a distortion attenuation.*

$$THD\,[\%] = \frac{\sqrt{V_2^2 + V_3^2 + \cdots + V_n^2}}{V_1} \times 100; \qquad THD\,[dB] = 20 \times \log(THD)$$

V_1 *– fundamental signal level [RMS, volts]*
V_2 *– second harmonic level [RMS, volts]*
V_n *– n^{th} harmonic level [RMS, volts]*

- **Total Harmonic Distortion + Noise (THD + Noise)**

  THD+N can be defined as the ratio of the square root of the sum of squares of the harmonic values (RMS) plus the square of the sum of all noise components and the square of the RMS level of the sinusoidal output signal.

$$THD + N = \frac{\sqrt{V_2^2 + V_3^2 + \cdots + V_n^2 + V_{Noise}^2}}{V_1}; \qquad THD + N\,[dB] = 20 \times \log(THD + N)$$

V_{Noise} *– noise level (RMS) within the measuring frequency bandwidth.*

In practice, it is common to measure THD + N:

THD+N is usually measured using a distortion analyzer.

The configuration of the measuring system is the same as in the previous measurements.

Measurement involves measuring THD + N at one frequency of the input signal to the amplifier.

Measurement procedure:

- Signal generator set to 1kHz sine wave and minimum output level.
- Amplifier volume control set to maximum.
- Carefully increase the output signal of the signal generator until the selected output power of the amplifier is reached, at which the THD + N measurement will be performed.
- Measure THD + N (%) according to the measurement procedure explained in the distortion analyzer user manual.
- Record the measurement result: THD + N = _ _ _ %, at f = 1kHz and P_{out} = _ _ _W
- Repeat the above measurement but using signal frequencies of 20 Hz, 40 Hz, 60 Hz, 80 Hz, 100Hz, 250 Hz, 500 Hz, 3 kHz, 5 kHz, 10 kHz, 12 kHz, 16 kHz and 20 kHz.

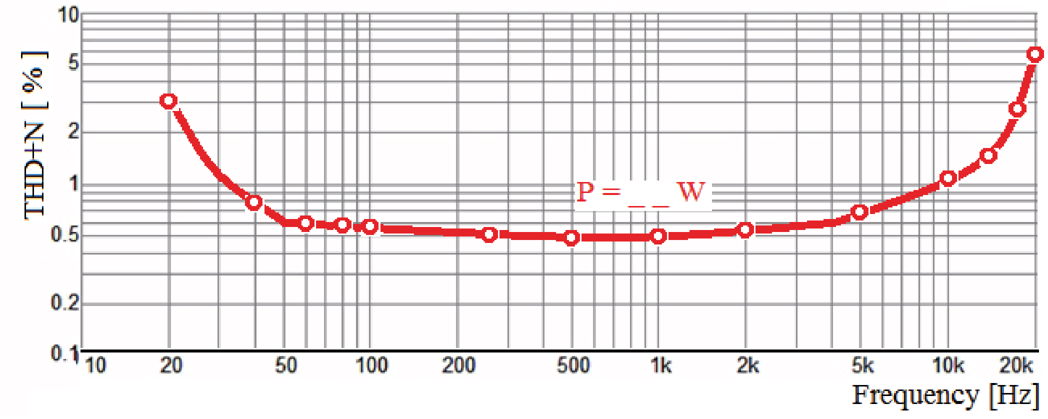
(Keep a constant output power during the measurement process).

P_{OUT} = _ _W	**20Hz**	40Hz	60Hz	80Hz	100Hz	250Hz	500Hz	**1kHz**
THD [%]								

P_{OUT} = _ _W	**1 kHz**	2 kHz	5 kHz	10 kHz	12 kHz	16 kHz	**20 kHz**
THD [%]							

The measurement results can be presented graphically and represent a mathematical function:

$$THD + N = f\,(frequency)\ \big|\,P_{out} = Constant$$

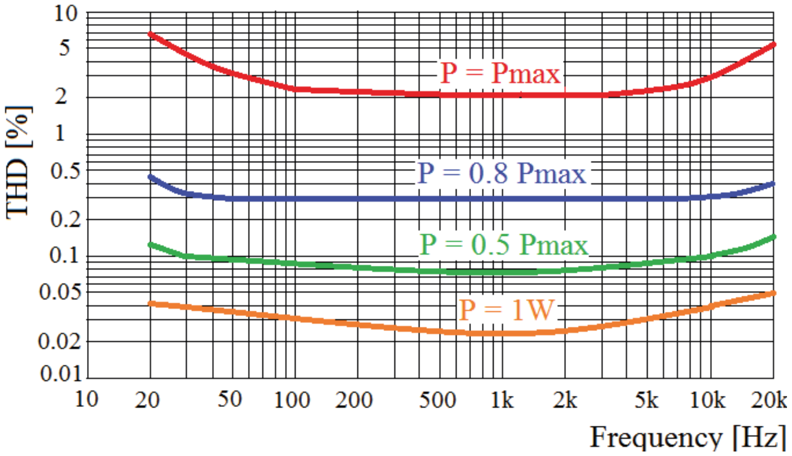


THD + N = f (frequency) | P_{out} *= Constant*
** In order to obtain a complete picture of the quality of the amplifier in terms of fidelity of the reproduction of the input signal, the measurement of THD + N is performed in the range of audible frequencies and at the several different values of the output power of the amplifier (from minimum to maximum).*

The complete previous measuring procedure of the THD+N of the amplifier should be repeated at the output power of the amplifier:

- $P_{out} = 0.8 \times P_{out\ (max)}$
- $P_{out} = 0.5 \times P_{out\ (max)}$
- $P_{out} = 0.25 \times P_{out\ (max)}$
- P_{out} = 1 W (R_L = 4 Ω, U_{out} = 2 V_{RMS}; R_L = 8 Ω, U_{out} = 2.83 V_{RMS})
- P_{out} = 250 mW (R_L = 4 Ω, U_{out} = 1 V_{RMS}; R_L = 8 Ω, U_{out} = 1.41 V_{RMS})

All sets of measured characteristics $THD + N = f\,(frequency)\ |P_{out} = Constant$ can be drawn in one common graph:



** Harmonic distortion depending on the frequency is generally plotted at different power levels in the audio spectrum.*

** Measurement of characteristics* $THD + N = f\,(frequency)\,|\ Pout = Ctc.$ *can be performed simultaneously with the measurement of the frequency response or the amplifier output power characteristics.*

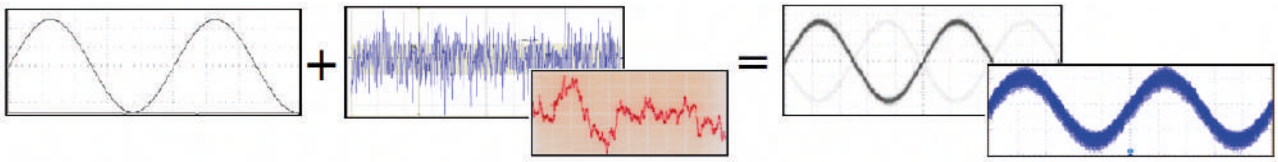
In order to get a more complete picture of the characteristics and quality of the amplifier, another set of measurements can be performed:

Signal to Noise Ratio (S / N)

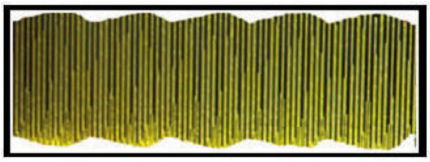
Signal to noise ratio is the ratio (commonly in dB) between the amplifier output signal level and the noise level at the output of the amplifier (noise generated by the amplifier itself or noise caused by external factors) when the input signal is set to zero.

- When defining S/N, the output signal level must be specified — i.e., output signal at the rated output power of the amplifier.

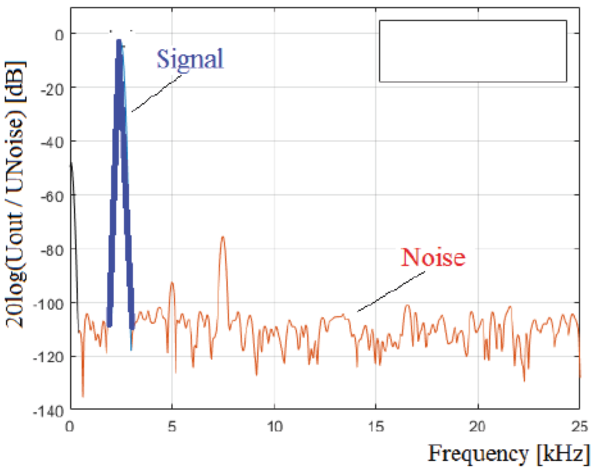
- Noise (or, self noise) is the sum of all noise (mainly thermal noise) generated by all electronics components of an audio amplifier. In the actual operating conditions of the amplifier, there are also unwanted signals coming from the mains supply and the power supply stage known as hum (or buzzing). There are also unwanted signals picked up from various external sources of electromagnetic field radiation.



Signal Noise



Output signal modulated with 100Hz hum signal caused by poorly filtered high voltage power supply (high ripple factor).



Measuring the S / N ratio of the amplifier is simply performed by measuring the rated level of the output signal of the amplifier (signal at rated output power) and measuring the level of the signal at the output of the amplifier without the presence of a signal at the input of the amplifier:

$$S/_N\,[dB] = 20 \times log\left(\frac{U_{OUT-rated}}{U_{Noise}}\right)$$

This means that the noise level is _ _dB below the signal level;

More precisely, in practice the ratio of signal plus noise and noise is measured:

$$(S+N)/_N\,[dB] = 20 \times log\left(\frac{U_{OUT-rated}+U_{Noise}}{U_{Noise}}\right)$$

Using the configuration of the measuring system as in the previous measurements (S + N) / N can be measured (more precisely, noise means noise plus hum).

Measurement procedure:

- Signal generator set to 1kHz sine wave and minimum output level.
- Amplifier volume control set to maximum.
- Carefully increase the output signal of the signal generator until the rated output power of the amplifier is reached and measure the output signal voltage ($U_{Out-RMS}$).
- Disconnect the input signal cable from the amplifier input signal connector. Short-circuit the input signal connector of the amplifier and measure the voltage at the output of the amplifier ($U_{Noise-RMS}$).

- Calculate:

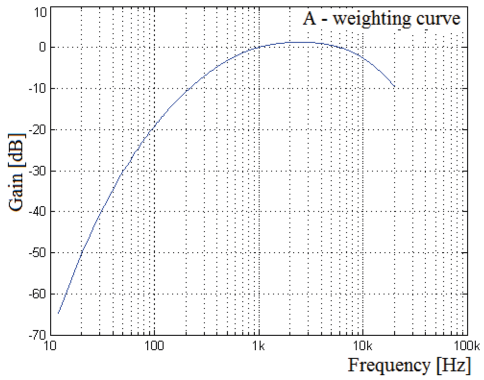
$$(S+N)/_N\,[dB] = 20 \times log\left(\frac{U_{Out-RMS}}{U_{Noise-RMS}}\right)$$

- If a decibel scale on an AC voltmeter is used, the measurement is significantly simplified: measure the dB level of the rated output signal level $[dB]_{OUT}$. Disconnect the input signal cable from the amplifier input signal connector. Short connect amplifier input signal connector and measure the signal level at the output of the amplifier $[dB]_{Noise}$.

  Calculate: **S / N [dB] = $[dB]_{OUT} - [dB]_{Noise}$**. (or, simply use the dB attenuator of the distortion analyzer AC voltmeter and the dB level of the output signal as a reference level).

* The measured value of noise voltage ($U_{Noise-RMS}$) at the output of the amplifier can be used as such when specifying the technical characteristics of the amplifier (usually used when specifying the technical characteristics of the preamplifiers): **Output noise: _ _ _ mV or μV**.

***** *S / N measurement using **A – weighting filter***

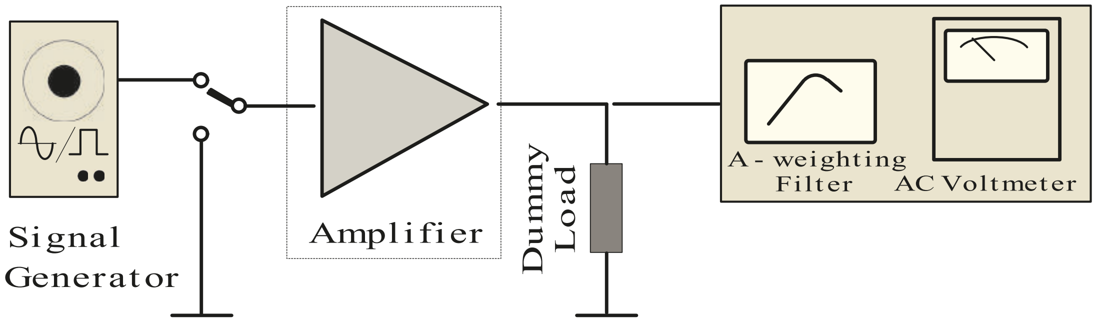


The reason of amplifier S / N measurement using A–weighting filter is to take into account the physiology of the human ear and the human perception of sound loudness at different frequencies within the audible frequency range described by the Fletcher – Munson curves (more simply, the A-weighting filter is used to simulate hearing sensitivity as a function of frequency).

A–weighting filter is a band–pass filter with a transfer characteristic that adjust the frequency spectrum of the signal to match the sensitivity of the human ear to a sound pressure level of 40 dB Fletcher – Munson curve.

("It weights the noise in accordance with human ear's perception of noise loudness").

Measurement system configuration
Use the same configuration as for S / N measurement explained above but with an A-weighting filter inserted between the amplifier output and the rest of the measuring equipment or use a measuring configuration using an AC voltmeter with a built-in A-weighting filter:



Measurement procedure:
Same as S /N measurement explained above.

The results of measuring the S / N ratio using an A-weighting filter are 10 to 20 dB better than the results of measuring without the use of A-weighting filter.

Example:
Hum and Noise (A – weighting at P_{Rated}) - _ _ _ dB

Measuring the S / N ratio of an amplifier provides useful information to the amplifier designer about the quality of the amplifier design and possible causes of poor S / N characteristic of the amplifier.

Some examples:

- By measuring and visually inspecting the noise and hum signals on the oscilloscope screen, an increased 50 Hz (60 Hz) hum signal was observed.
  Possible causes:
  - *Layout of electronic components (electrostatic and electromagnetic couplings).*
  - *Orientation and location of mains power supply transformer to other electronic components.*
  - *Mains power supply transformer is located close to the input stage of the amplifier.*
  - *Wires connecting the primary winding of the mains transformer to the mains connector (power switch, fuse, RSO filter) or wires connecting the secondary windings to the rectifier stage placed close to the components and wires of the useful signal path.*
  - *AC tube heating (unwanted humming can be minimized by using a balanced AC heater power supply circuit or by using a DC heater power supply).*

- 100 Hz (120 Hz) hum (manifests as modulation of the wanted signal by a 100 Hz (120 Hz) signal.
  Possible causes:
  - *High Ripple Factor (ripple voltage) of the high voltage power supply in general and unacceptable ripple factor of the high voltage power supply of each amplifier stage (input, driver, output).*
  - *Grounding – grounding technique used, grounding loops, insufficient (small) cross section of grounding wire.*

- High frequency noise
  Possible causes:
  - *Used high-noise electronic components (tubes, resistors).*
  - *Long wiring along the signal path (keep the signal path as short as it is possible).*
  - *High impedance of the input circuit of the tube of each stage of the amplifier.*
  - *Unshielded cable used in signal path wiring.*
  - *Inadequate NFB frequency compensation (amplifier with NFB).*

After all the measurements described above, the technical specification of the audio amplifiers can be defined:

- Single End class A power amplifier (or, Push Pull class AB (A)).
- Tubes: ___ , ___ , ___ , ____ ,
- Amplification: A = _ _ _ [×], A = _ _ _ [dB]
- Load: 4Ω, 8Ω.
- Input Impedance: _ _ _ kΩ.
- Rated Output Power: P_{Out} = _ _ _ *W, into _ _ _ Ω at THD +N ≤ _ _ _ % [20Hz ÷ 20kHz].*
- Input Sensitivity: U_{in} = _ _ _ mV, for P_{out} rated
- Frequency Response (- 3dB): [_ _ _ Hz ÷ _ _ _ kHz] , at rated P_{out} .
- Frequency response: (_ _ _ dB at 20Hz, _ _ _ dB at 20 kHz), at rated. P_{out}
- Frequency Response (- 3dB): [_ _ _ Hz ÷ _ _ _ kHz] , at P_{out} = 1 W .
- Power Bandwidth: (_ _ _ Hz ÷ _ _ _ kHz).
- Total Harmonic Distortion: THD + N ≤ _ _ _ % [20 Hz ÷ 20 kHz] at rated P_{out}.
- Total Harmonic Distortion: THD + N ≤ _ _ _ % [20 Hz ÷ 20 kHz] at P_{out} = 1W.
- Total Harmonic Distortion: THD + N ≤ _ _ _ % [1 kHz] at rated P_{out} = ___ .
- Signal to Noise Ratio: S + N / N ≥ _ _ _ dB, at rated P_{out}.
- Hum and Noise (A – weighting at rated P_{out}) : _ _ _ dB
- Maximum Output Power: P_{max} = _ _ _ W, at 1 kHz.

Power Input: _ _ _ VAC (+10, −15)%, 50 Hz, or : _ _ _ VAC 60Hz
Rated Power Consumption: _ _ _ W, or _ _ _ VA
Size (W × H × D): (_ _ _ × _ _ _ × _ _ _) mm
Weight: _ _ _ kg

*** Preamplifier performance measurement**

Measuring the performance of a preamplifier is similar to measuring the performance of a power amplifiers and uses the same measuring equipment and measurement system configuration.
As with power amplifiers, it is necessary to perform both DC and AC measurements of the preamplifier.

- AC measurement
  * Output power-related characteristics are not measured.

  - **Frequency Response** or Amplitude Characteristic is one of the most important characteristics of a preamplifier.

    Amplification expressed as the ratio of output to input RMS voltage ($U_{out\ -\ RMS}$ / $U_{in\ -\ RMS}$) or [20 × log ($U_{out\ -\ RMS}$ / $U_{in\ -\ RMS}$)] dB at **1 kHz** input signal is the **reference level (U** **U_{REF})**.
    ($U_{out\ -RMS}$ is usually measured at the output of the preamplifier terminated with load resistance of 100 kΩ).

Perform measurements (keep the input signal level constant):

$$A = \frac{U_{out}}{U_{in}} = f(frequency)\Big|_{U_{in=ctc}} \quad or,$$

$$A(dB) = 20 \times log\frac{U_{out}}{U_{in}} = f(frequency)\Big|_{U_{in=ctc}}$$

Record the measurement results (present graphically).

Specify:

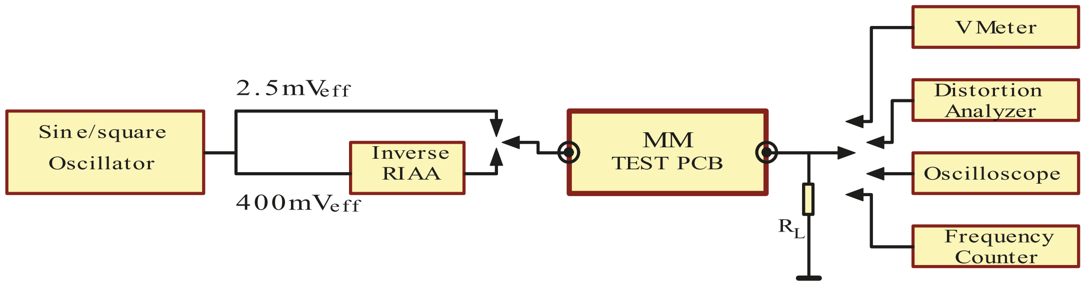
Amplification: _ _ _ dB at 1 kHz.

Frequency response: [f_L ÷ f_H], (−3 dB amplitude frequency band) at U_{out} = _ _ _ mV.

Frequency response: _ _ _ dB at 20 Hz and _ _ _ dB at 20 kHz.

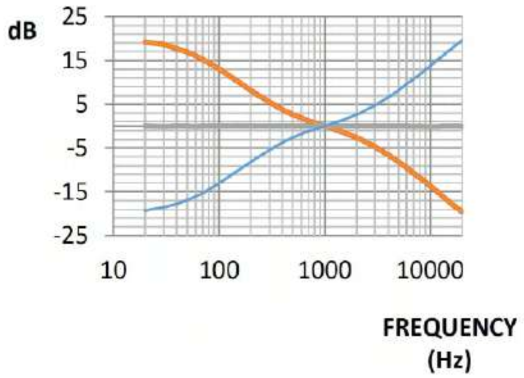
** Measurement of the frequency response of a preamplifier with specific transfer characteristic can be perform in a standard way by measuring U_{out} = f (frequency or using specific filters. A typical example is the RIAA preamplifier. Measurement of the transfer characteristic or amplitude response of the RIAA preamplifier in a standard way can be very difficult because the amplitude of the output signal varies in a range of ± 20 dB in the frequency range of 20 Hz to 2 0kHz. Due to the limited measuring ranges of the most commonly used AC voltmeters, measuring can be very inconvenient and can affect the accuracy of the measurement. Therefore, the inverse RIAA filter is more commonly used to measure the frequency response of RIAA preamplifier.*
The following information (data) must be provided in the technical specification of the RIAA preamplifier: deviation of the amplitude characteristic of the preamplifier from the standard RIAA characteristic.

Example:
MM RIAA preamplifier
Measurement of amplitude characteristic
Measurement system configuration:



1. Standard method of measurement

Frequency [Hz]	U out [mV]	A [dB]	DEVIATION Δ [dB]
20	9233.7	19.221	- 0.053
30	8603.5	18.607	0.014
40	7755	17.705	- 0.087
50	7093	16.930	- 0.016
70	5809	15.195	- 0.087
100	4545	13.064	- 0.025
200	2590	8.179	- 0.04
300	1892.5	5.454	- 0.03
400	1556.7	3.757	- 0.026
500	1364.5	2.613	- 0.035
800	1100	0.741	- 0.008
1000	**1010**	**0**	**0**
2000	749	-2.596	- 0.008
2122	719	- 2.952	- 0.085
3000	584.6	-4.749	- 0.009
5000	392.5	-8.209	0.001
8000	256.7	-11.898	- 0.004
10000	208	-13.725	0.009
11000	192.5	-14.502	0.028
12000	175	-15.225	0.034
15000	140	-17.163	- 0.007
16000	131	-17.741	- 0.031
17000	123.5	-18.253	- 0.023
18000	116.5	-18.760	-0.04
19000	111	-19.180	0
20000	106	-19.580	0.04



The measurement of the amplitude characteristic is performed using a signal of constant level (approximately 2.5 mV$_{eff}$) and variable frequency in the range from 20 Hz to 20 kHz. By calculating the ratio of the measured values of the signal voltage at the output of the preamplifier at different frequencies and the voltage of the signal frequency of 1kHz as the reference level at the output of the preamplifier, the amplitude characteristic of the preamplifier is determined:

$$A = \frac{U_{out}}{U_{[1000Hz]}} = \frac{U_{out}}{1010mV}; \quad A[dB] = 20 \times \log\left(\frac{U_{out}}{U_{[1000Hz]}}\right)$$

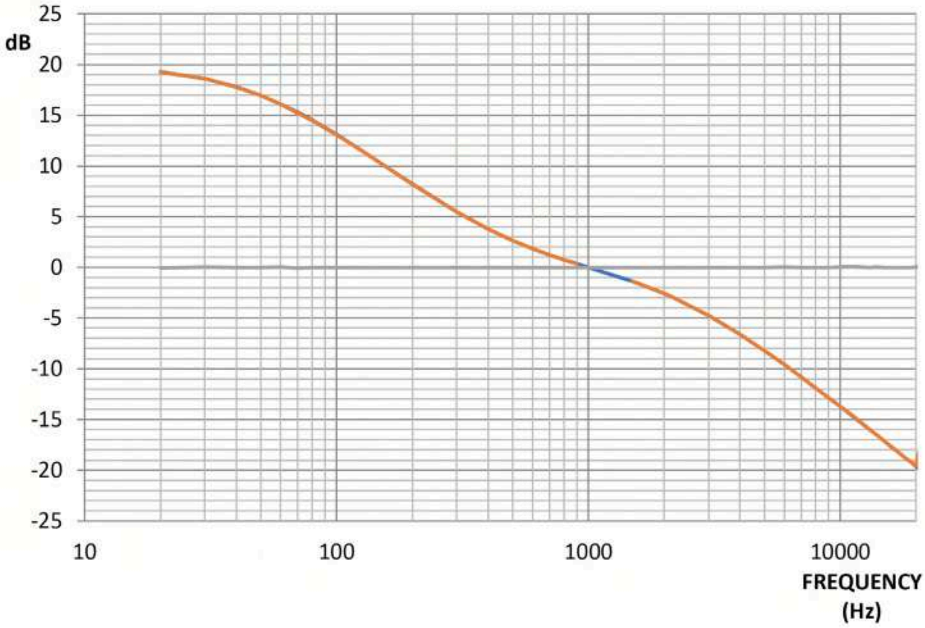
$$A[dB] = 20 \times \log\left(\frac{U_{out}}{1010mV}\right)$$

$$\Delta[\mathrm{dB}] = \mathrm{A}[\mathrm{dB}]_{\mathrm{RIAA\ standard}} - \mathrm{A}[\mathrm{dB}]$$

Note:

The measured values of the output signal level are in the range from 106 mV$_{eff}$ to 9.2 V$_{eff}$ and cannot be measured using only one measuring range of the AC voltmeter.

Graph of the amplitude characteristic of the MM RIAA preamplifier (with some corrections and interpolation):

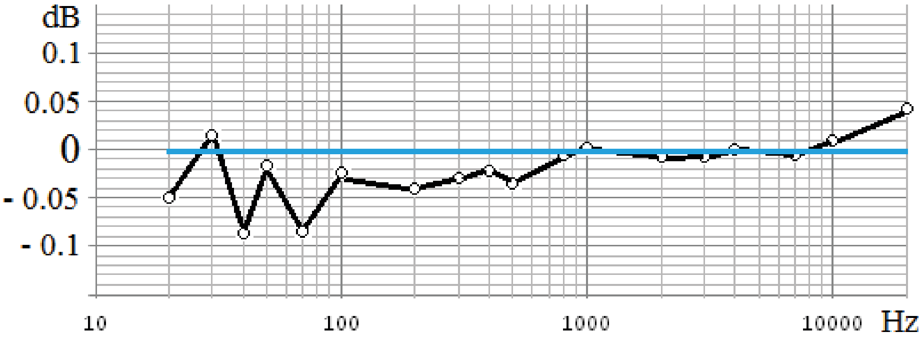


2. Measurement method using an inverse RIAA filter:

The measurement results using this method represent the deviation of the amplitude characteristic of the MM RIAA preamplifier from the standard RIAA characteristic.

Graphical representation:

$$\Delta[\mathrm{dB}] = A[\mathrm{dB}]_{\mathrm{RIAA\ standard}} - A[\mathrm{dB}]$$



Total Harmonic Distortion (THD) or
Total harmonic Distortion + Noise (THD+N)

- Signal generator set to 1 kHz sine wave and minimum output level.
- Preamplifier volume control (if preamplifier has it) set to maximum.
- Carefully increase the level of the output signal from the signal generator and visually monitor the image of the output signal of the preamplifier on the oscilloscope screen.
- Increase the output signal from the signal generator until it is visible on the oscilloscope screen that the sine wave signal has started to clip.
- Slightly reduce the input signal to the preamplifier until the sine wave on the oscilloscope screen returns to the shape of an undistorted sine wave.
- Read and record the RMS voltage at the output of the preamplifier on the AC voltmeter of the distortion analyzer – **maximum output voltage $U_{out - max}$.**
  (Read and record the RMS voltage at the input of the preamplifier, also –**maximum input voltage $U_{in - max}$**).
- Measure THD +N. (THD: _ _ _ %, at $U_{out - max}$, and 1kHz).
- Measure THD + N at **rated input signal level** (**$U_{in\ (rated)}$**) and frequencies: 20 Hz, 40 Hz, 60 Hz, 80Hz, 100 Hz, 250 Hz, 500 Hz, 3 kHz, 5 kHz, 10 kHz, 12 kHz, 16 kHz and 20 kHz. (Keep the input signal level constant).
- Record the measurement result (display graphically).

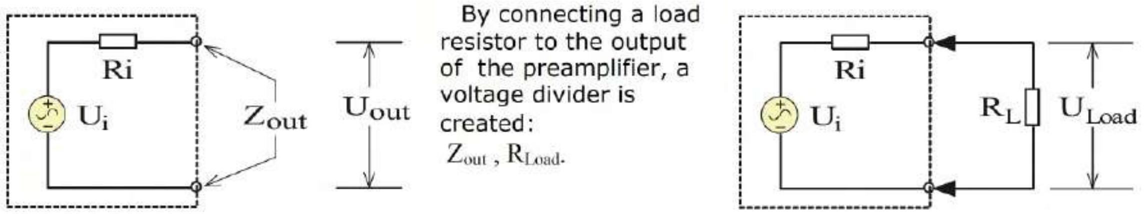
Signal to Noise Ratio (S / N)
Same procedure as for measuring S / N power amplifier.

Output Impedance
Preamplifier output impedance is a very important characteristic, as well as other characteristics such as, for example, the level of the output signal, because it represents the ability of the preamplifier to drive the power amplifier in a qualitative way. The output impedance of the preamplifier is a frequency-dependent quantity but is usually specified at 1kHz.

Therefore, the measurement of the output impedance of the preamplifiers should involve a 1-kHz input signal and the appropriate level.

The most common and simplest method of measuring the output impedance of a preamplifier is based on the application of basic laws of electric current to the voltage divider consisting of the output impedance of the preamplifier and the resistance of external load connected to the preamplifier output.

A device whose output impedance is measured can be presented as an output signal generator with its own internal resistance:

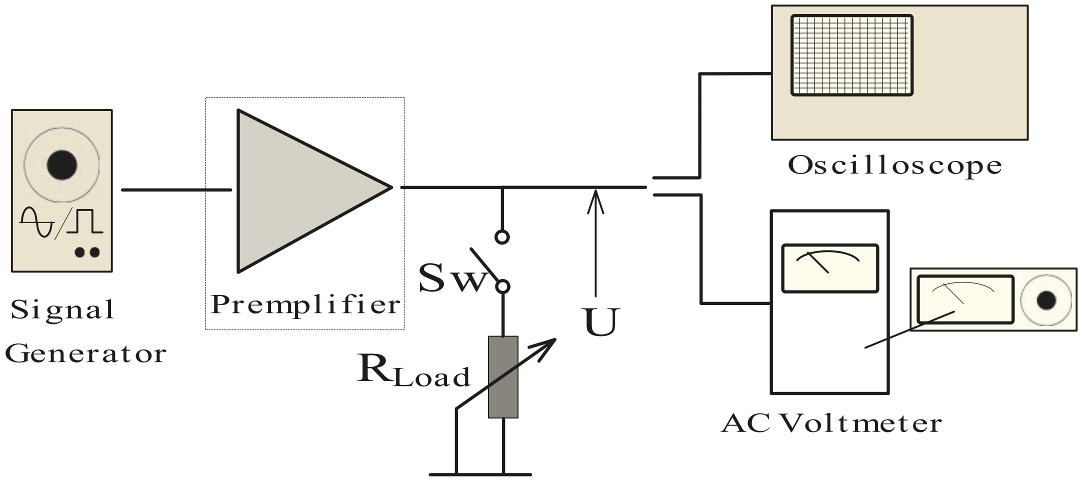


By applying the Kirchhoff's laws, the output impedance is:

$$Z_{out} = \frac{U_{out} - U_{Load}}{\frac{U_{Load}}{R_{Load}}} = R_{Load}\left(\frac{U_{out}}{U_{Load}} - 1\right)$$

A special case: $U_L = \frac{U_{out}}{2} \rightarrow \boldsymbol{R_L = R_{out}}(Z_{out})$; (it is used in a so-called „half-voltage" method of measurement).

Measurement procedure:
The configuration of the measuring system is very simple:



- Signal generator set to 1kHz sine wave.
- Measure the voltage at the unloaded output of the preamplifier $\mathbf{U_{out-0}}$ (or, dB output level) and record it.
- Connect the variable resistor R_{Load} to the output of the preamplifier. Set the variable resistor to its maximum value.
- Carefully reduce the R_{Load} resistance until half the voltage (or, –6 dB) of the unloaded preamplifier is measured on the AC voltmeter: $\mathbf{U_{Load} = U_{out-0} / 2}$.
- According the Kirchhoff's laws, the output impedance is:

$$Z_{out} = \frac{U_{out} - U_{Load}}{\frac{U_{Load}}{R_{Load}}} = R_{Load}\left(\frac{U_{out}}{U_{Load}} - 1\right)$$

  A special case:

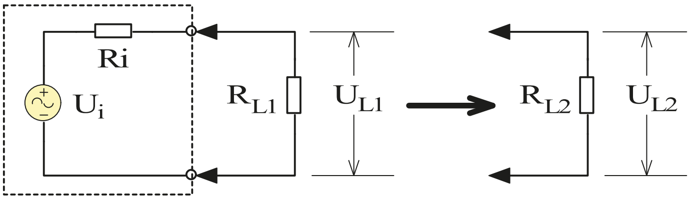
$$U_L = \frac{U_{out}}{2} \rightarrow R_L = R_{out}(Z_{out})$$

- Disconnect the variable R_{Load} resistor from the preamplifier output and measure its resistance using a high-precision ohmmeter. **The measured resistance is equal to the preamplifier output impedance**.

** The choice of load resistance should be consistent with the fact that the output impedance of the preamplifier ranges from a few hundred ohms to several kΩ in most cases.*
When measuring, it is necessary to select (or re adjust) the level of the input signal so that when the output of the preamplifier is loaded, an undistorted sine wave is obtained (monitoring on oscilloscope screen).

Another variant of the above method of measuring the output impedance of the preamplifier

By measuring the voltage at the preamp output at two different load resistances:



Measured voltage values at loads:
- R_{L1}: U_{L1}
- R_{L2} : U_{L2}

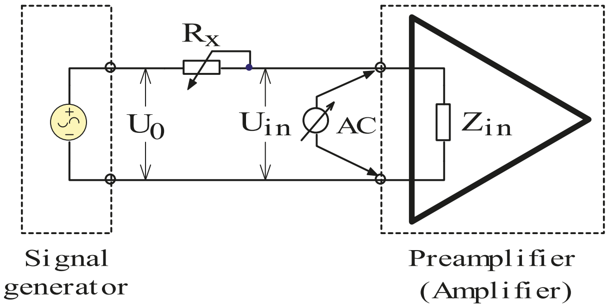
Calculate the output impedance of the preamplifier using the equation:

$$R_{out}(Z_{out}) = \frac{R_{L1} - R_{L1} \times \frac{U_{L1}}{U_{L2}}}{\frac{U_{L1}}{U_{L2}} - \frac{R_{L1}}{R_{L2}}} = \frac{R_{L1}\left(1 - \frac{U_{L1}}{U_{L2}}\right)}{\frac{U_{L1}}{U_{L2}} - \frac{R_{L1}}{R_{L2}}}$$

** Note*
- *For both of the above measurement procedures, use lower levels of the input signal to avoid problems related to distortion of the output signal of the loaded preamplifier (amplifier).*
- *Both of the above methods can be used to measure the output impedance of a power amplifier (the load resistor used must be of adequate power depending on the output power of the amplifier).*

Input Impedance
The input impedance of an (amplifier) is the input impedance "seen" by the source driving the input of the *preamplifier (amplifier).*

Measurement of the input impedance of the preamplifier is based on the same theoretical principles as the measurement of the output impedance of the preamplifier.
The practical realization of the measurement is performed by inserting a variable resistor between the signal generator and the input of the preamplifier (voltage divider is created: Z_{in} , R_x):



According the Kirchhoff's laws, the input impedance is:

$$Z_{in} = R_x \times \frac{U_{in}}{U_0 - U_{in}}$$

A special case: ;

$$U_{in} = \frac{U_0}{2} \rightarrow \boldsymbol{R_x} = \boldsymbol{R_{in}}(Z_{in});$$

The simplest way to measure the input impedance of a preamplifier is to use the so-called "half-voltage" method.

** Although the input impedance of the preamplifier depends on the frequency of the input signal, it is common for the input impedance to be measured and specified at a signal frequency of 1 kHz.*

Measurement procedure:

- Signal generator set to a sine wave signal frequency of 1 kHz and a level equal (or lower) to the rated input voltage of the preamplifier (**U_0)**.
- Connect the variable resistor R_x between the signal generator and the input of the preamplifier. Set the variable resistor to its minimum value.
- Connect an AC voltmeter to the input of the preamplifier.
- Measure the voltage at the input of the preamplifier **U_0** (or dB) and record it.
- Carefully increase the resistance R_x until the voltage at the preamplifier input drops to a value equal to half (**U_0 / 2)** the voltage of the signal generator (or, –6dB) **(**measured by the AC voltmeter).

- According the Kirchhoff's laws, the input impedance is:

$$Z_{in} = R_x \times \frac{U_{in}}{U_0 - U_{in}}$$

A special case: $U_{in} = \frac{U_0}{2} \rightarrow \boldsymbol{R_x} = \boldsymbol{R_{in}}(Z_{in})$;

- Disconnect the variable R_x resistor from the circuit and measure its resistance using a high-precision ohmmeter. **The measured resistance is equal to the preamplifier input impedance**.

Preamplifier Technical Specification

- Amplification: A = _ _ _ [×], A = _ _ _ [dB]
- Maximum Input signal: $U_{in\text{ - }max}$ = _ _ _ mV, ($U_{out - max}$ = _ _ _ V_{RMS}, 1 kHz)
- Input Impedance: _ _ _ kΩ.
- Input Sensitivity: U_{in} = _ _ _ mV, at rated U_{out}
- Output Impedance: _ _ _ Ω.
- Frequency Response (–3dB): [_ _ _ Hz ÷ _ _ _ kHz] , at rated U_{out}.
- Frequency response: (_ _ _ dB at 20Hz, _ _ _ dB at 20 kHz), at rated U_{out}.
- Total Harmonic Distortion: THD + N ≤ _ _ _ % [20 Hz to 20 kHz] at rated U_{out}.
- Signal to Noise Ratio: S + N / N ≥ _ _ _ dB, at rated U_{out}.
- Hum and Noise (A–weighting at rated U_{out}) : _ _ _ dB.

MM RIAA Preamplifier

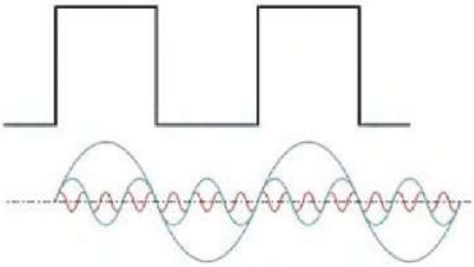
- Amplification: A = _ _ _ [×], A = _ _ _ [dB] at 1kHz.
- Input Impedance: 47 kΩ.
- Input Sensitivity: U_{in} = 2.5 mV
- Deviation from the RIAA standard curve: ± _ _ _ dB , [20 Hz ÷ 20 kHz]
- Output Impedance: _ _ _ Ω.
- Total Harmonic Distortion: THD + N ≤ _ _ _ % [20 Hz ÷ 20 kHz] at rated U_{out}
- Signal to Noise Ratio: S + N / N ≥ _ _ _ dB, at rated U_{out}.
- Hum and Noise (A–weighting at rated U_{out}) : _ _ _ dB.

Square Wave Testing

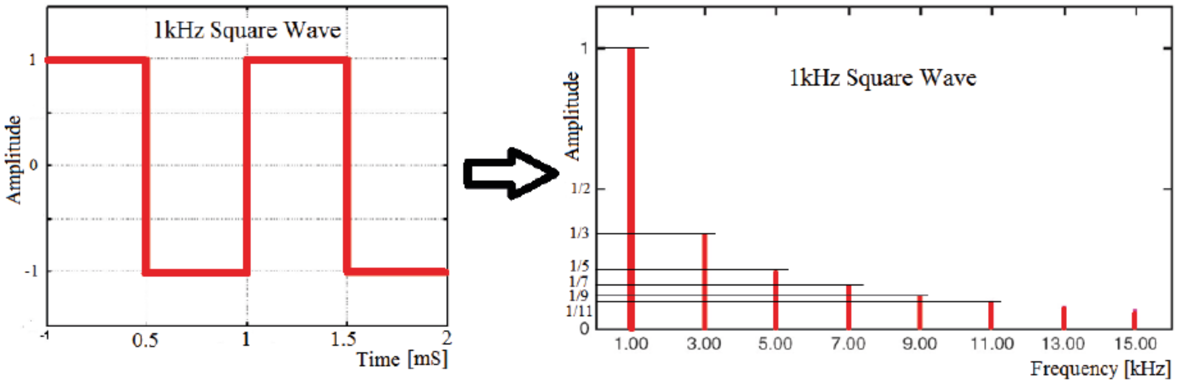
Most audio signal generators, in addition to generating a sine wave signal, also generate a square signal. Why is a square wave signal used in audio amplifier testing? A mathematical operation known as the Fourier transform showed that any waveform could be modeled as a combination of different types of sine waves. If it is applied to a square wave signal, the result is that the square wave can be represented as an infinite sum of sinusoidal waves. Square waves produce an infinite number of the harmonics, but only odd integer ones. The amplitude of each harmonic is equal to the reciprocal value of its harmonic number (the third harmonic has amplitude of 1/3 of the amplitude of the signal at the fundamental frequency, the fifth: 1/5, the seventh: 1/7, and so on).

The square wave can be represented by the equation:

$$x(t) = A \times \left(sin(\omega t) + \frac{1}{3} sin(3\omega t) + \frac{1}{5} sin(5\omega t) + \cdots \right)$$



Plot of the spectrum of a square wave:



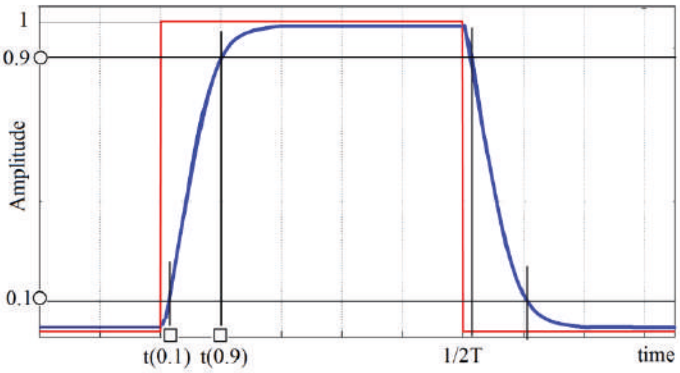
- The RMS value of the square wave signal:

  $\mathbf{U_{RMS} = U_{peak}}$

- Harmonics:

  $U_n = \frac{2}{(\pi \times n)} \times U_{p-p}$; n – number of harmonic (odd integer)

Slew Rate:



Rate of change from 10% to 90% of the amplitude of the square wave (change of voltage per unit of time).

$$Slew\ Rate\ (S) = \frac{dU_0}{dt} = \frac{U_{0\,(0.9)} - U_{0\,(0.1)}}{t_{(0.9)} - t_{(0.1)}} \left[\frac{V}{s}\right]$$

As the square wave signal is rich in integer odd harmonics, this qualifies it as a powerful tool for analyzing the frequency response of amplifiers.

Test procedure:
Measurement system configuration — the same as for basic AC measurement.
The square wave function of the signal generator is used.

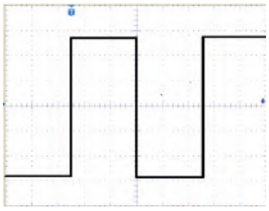
In practice, this method of testing amplifiers cannot give quantifiable measurement results taking into account the simple measuring equipment used (for example, a 10-MHz analog oscilloscope), their resolution and accuracy.
For these reasons, this method of testing is based on human vision as the main testing instrument, as well as on the skills, technical knowledge and experience of the person performing the testing.

However, this method of testing is subjective and depends on an imperfect test instrument — the human eye, but careful observation of the shape of the output signal of the amplifier on the oscilloscope screen can provide very useful data on the frequency response of amplifiers and possible design problems.

Testing should begin with a square wave signal of medium frequencies (middle frequency range), i.e. using a 1-kHz square wave signal.

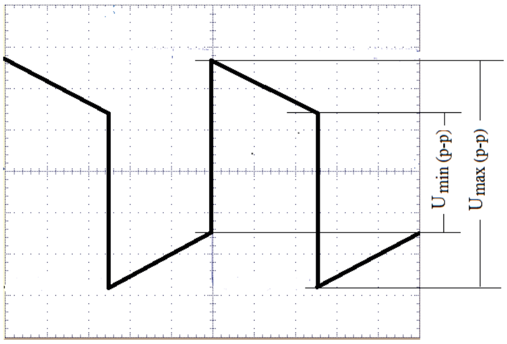
Depending on the technical characteristics of the amplifier (transfer characteristic of the amplifier), different shapes of the square wave signal (output signal of the amplifier) can be seen on the oscilloscope screen:

a) 1-kHz Square wave



An almost ideal square signal shape.
Conclusion at first sight:
The amplifier seems to have a good frequency response (wide frequency range) and no design problems.

b) 1kHz Square wave

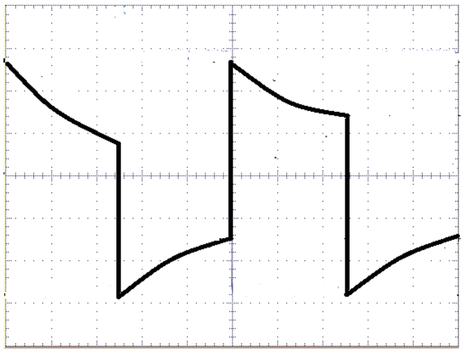


The normally flat top is sloping.
Conclusion at first sight: The frequency response of the amplifier at low frequencies is poor (**low frequencies are attenuated**).
Probably unacceptably **high** Cut-off Low Frequency f_L (-3dB).
The approximate Cut-off Low Frequency f_L (–3 dB) can be calculated from:

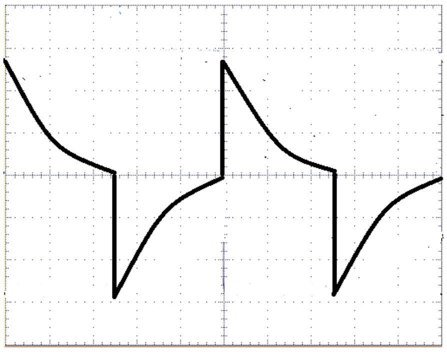
$$f_L = \frac{\left(U_{\max(p-p)} - U_{\min(p-p)}\right) \times f}{\pi \times U_{\max(p-p)}}$$

f – frequency of the square wave signal [Hz]

If the frequency of the signal decreases or shifts to f_L, the shape of the signal on the oscilloscope screen also changes:



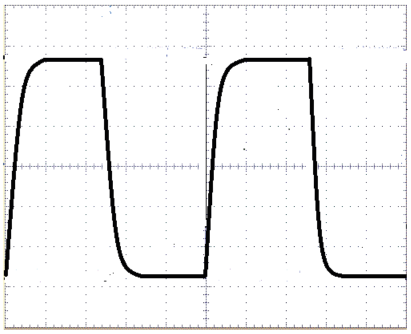
f_{signal} < 1 kHz Square wave



Limiting case: $f_{signal} = f_L$

*The possible cause of the high cut-off low frequency is the **low (insufficient) capacitance of the coupling capacitors**, for example or poor quality of output transformer**.***

c) 1-kHz Square wave

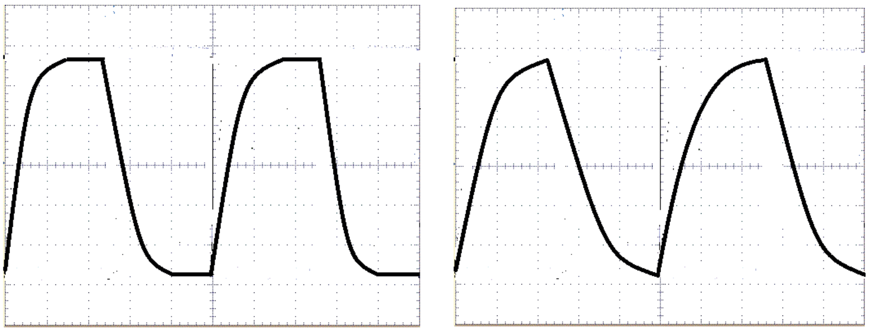


The square wave is rounded.
Conclusion at first sight: The amplifier frequency response at high frequencies is bad (**high frequencies are attenuated**).

Probably unacceptably **low** Cut-off High Frequency f_H (–3 dB).
The approximate Low Cutoff Frequency f_L (-3dB) can be calculated by the equation:

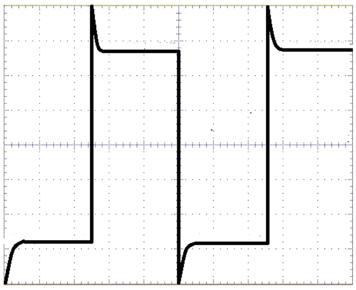
$$f_H = \frac{0.35}{t_{(0.9)} - t_{(0.1)}}$$

If the frequency of the signal increases or shifts to f_H, the shape of the signal on the oscilloscope screen also changes:



f_{signal} > 1 kHz Square wave *Limiting case: $f_{signal} = f_H$*

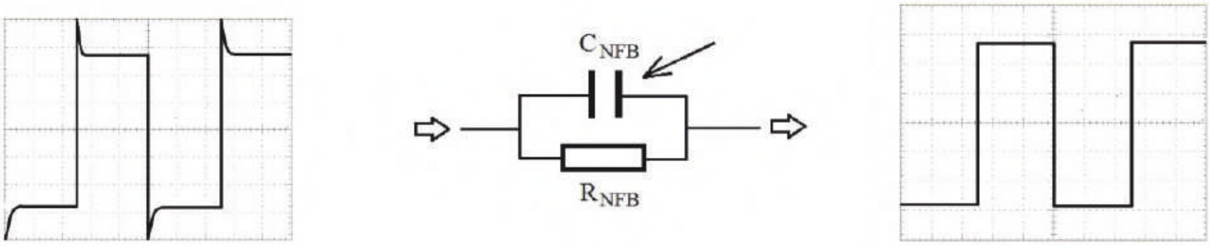
d) 1 kHz Square wave



The leading edge of the square wave extends beyond the flat part of the square wave curve ("overshoot").

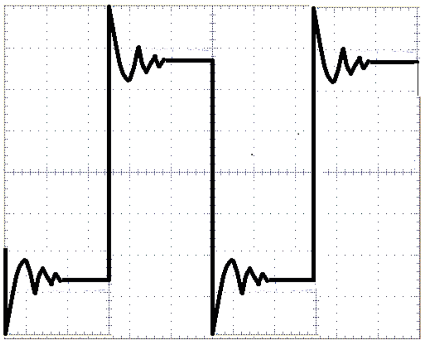
Conclusion at first sight: increased amplification at high frequencies.

Example:
For amplifiers with applied NFB, it is necessary to fine-tune the frequency-dependent NFB circuit. By choosing the correct capacitance in the NFB circuit, an almost ideal square wave shape can be achieved:



* *Check the shape of the square signal at frequencies of 10 kHz and 20 kHz (also, 20 Hz, 100 Hz).*
Testing amplifier using the square signal method can be of great benefit to the designer when adjusting the NFB of the amplifier.

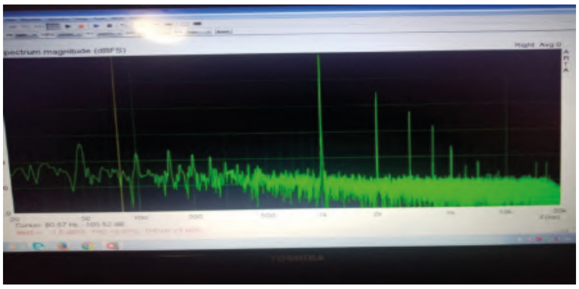
e) 1 kHz Square wave



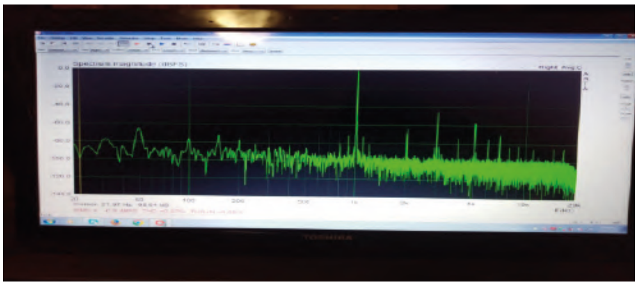
Relaxation oscillations at the top of the curve – **ringing**.
Conclusion at first sight: the existence of resonant circuits inside the amplifiers – the cause of ringing at high frequency.

It is very difficult to determine the cause of this problem.
One of the causes may be the resonant frequency of the output transformer (inductance and capacitance of the primary winding of the output transformer form an oscillatory circuit).

The use of a Zobel filter is preferred.



Single-Ended amplifier FFT analysis



Push-Pull amplifier FFT analysis

Other specific characteristics of amplifiers and preamplifiers can be measured using advanced measurement methods (FFT – Fast Fourier Transform) and sophisticated measuring devices (spectrum analyzers). FFT analysis is a very useful method used in the analysis of amplifier and preamplifier performance. In the measurement methods described so far, the treated analog signals are presented in the time domain and their physical measurement can be performed using analog instruments. FFT analysis is not actually a physical measurement of an analog signal, but it is in fact a mathematical algorithm (computer program) that transforms a signal from a time domain into a frequency domain. FFT analysis obtained a decomposed analog signal in the frequency domain, that is, a visual representation of the fundamental harmonic and all other harmonics contained in the treated signal - their frequencies and amplitudes. FFT analysis also provides other very useful data on the quality of amplifiers such as THD, THD + N, noise, S / N, ... For example, FFT analysis is very useful in designing and measuring the power supply stage of an amplifier or in the AC balancing process of a push pull amplifier. Today, in the age of personal computers and the Internet, free sound analysis software can easily be found.

At home, satisfactory measurements can be obtained by using the classical measurement methods described above and by using simpler measuring devices.
Example of a typical set of measuring equipment for measuring the performance of amplifiers and preamplifiers at home:

- Sine / Square Oscillator 10 Hz to 1 MHz, PM 5126 Philips
- Universal frequency counter 120 MHz, PM 6669 Philips
- Distortion Analyzer 334 A, Hewlett Packard
- Oscilloscope 0 to 15MHz, PM 3207 Philips
- Digital VAΩ meter, PM 2522 Philips
- 3½ digit multimeter DT 9208 A Yihai Int'l Group
- 3½ digit Capacitance Meter CM 9601A
- Dummy load 100 kΩ / 2 W Metal Film resistor
- Dummy load 4 Ω, 8 Ω, / 100 W
- Inverse RIAA filter (–44 dB)
- Autotransformer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Index

ELEKTOR ACADEMY PRO

High-End Tube Amplifier Design

A Toolbox for Audio Lovers and Engineers

Without any ambition to reach scientific levels, this book aims to be a toolbox for both audio lovers and high-end equipment designers. The elementary theory presented is the bare minimum for readers to grasp the operation and practical use of electrical, electromagnetic, physics, and electronic operations available in the designers' toolbox. Each tool is explained in a minimum of words and theory without needless coverage of underlying equations or figures.

The book chapters guide you through the process of designing quality amplifiers with vacuum tubes, from the very beginning, considering both technical and subjective requirements — in theory and practice.

The book is a compilation of the author's notes used in his professional and educational career but was nevertheless primarily written as a result of true love for the audiophile hobby.

Zoran M. Dukic was born in Kraljevo, Serbia in 1953. He was Winner of the Yugoslavian National High School Physics Competition in 1971 and graduated in 1972. In 1978, he graduated from the Faculty of Electronics at the University of Nis. Zoran began his professional career at the Ei Radio Tube Factory, continued at the Ei Radio Acoustics Factory, and ended as Deputy General Director of Ei Holding Corporation. He is an external expert associate at the Faculty of Electronics in Nis (quantum and optical electronics).

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