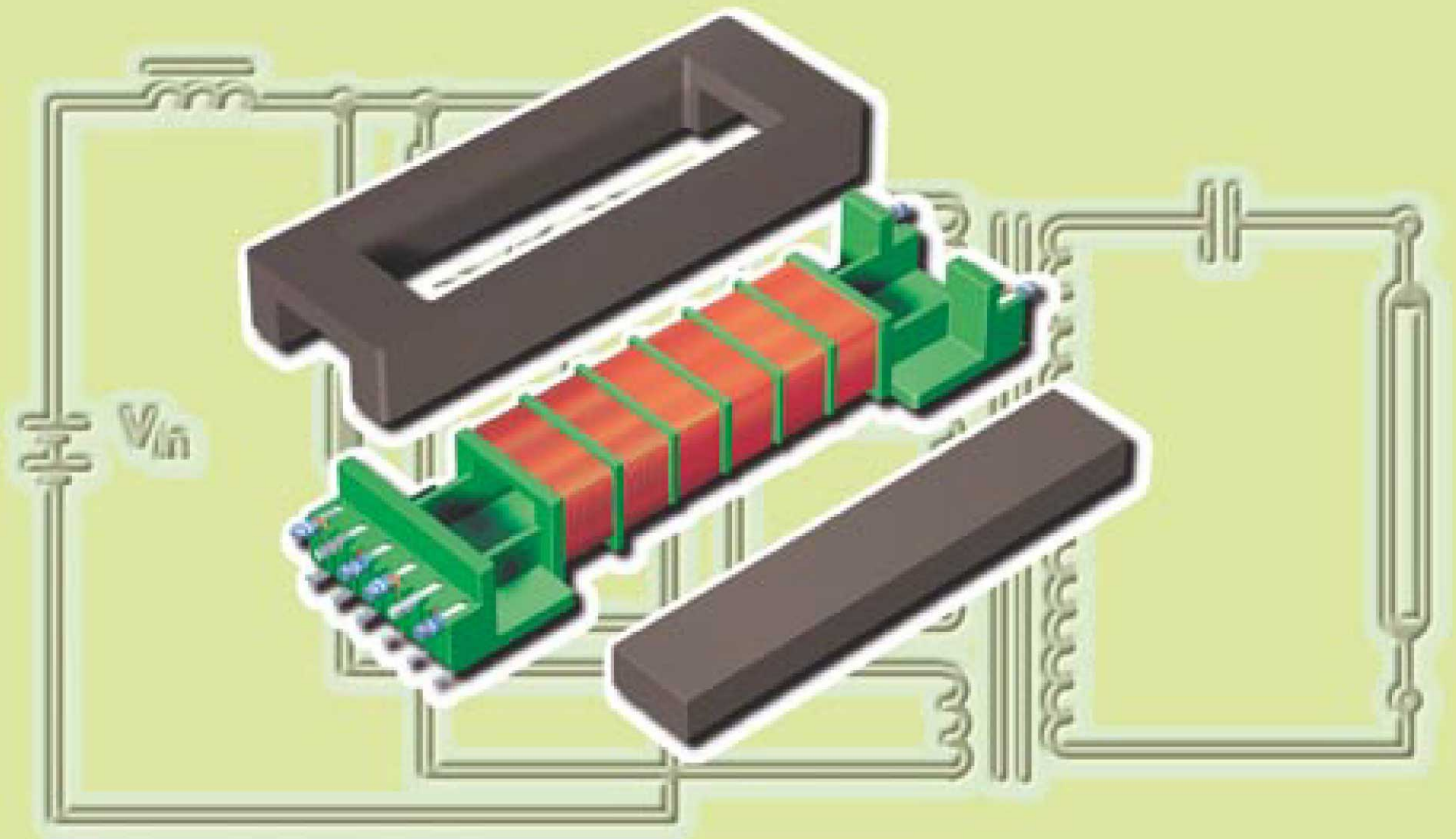


Design of CCFL Backlight Inverters with Frame & Bar Cores



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Summary

Portable devices, such as notebook computers and personal digital assistants are developed rapidly nowadays, which places an increasing demand on display technology. The LCD with cold cathode fluorescent back lighting satisfies the requirements on display performance, size and efficiency.

The backlight inverter plays a crucial role in supplying the power to the lamps of these flat-panel displays.

The frame and bar ferrite cores, described in this brochure, are especially designed to build the slender, low profile transformers required for these inverters.

A worked out design example is given, as well as data sheets of the standard range of 4 sizes frame and bar cores in 3C90 and 3C91, including the dedicated SMD coil formers.

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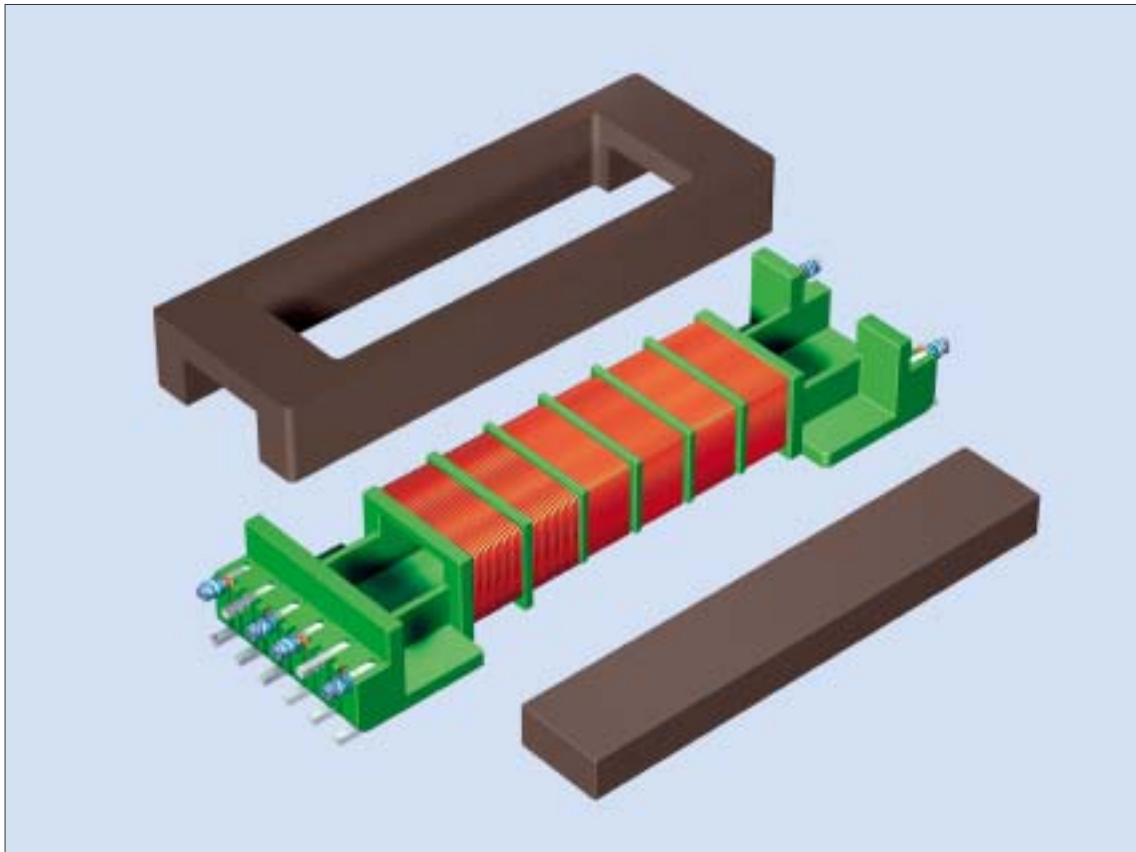


Fig.1 Exploded view of a Frame and Bar core assembly with wired bobbin

I. Introduction

Portable devices, such as notebook computers and personal digital assistants (PDA) are developed rapidly nowadays, which places an increasing demand on display technology. Notebooks use very powerful and thus energy consuming processors; the processor consumes up to 60% while the backlight inverter needs only 10 to 20% of the battery power. The tasks of a PDA are less and so are the requirements of its processor. In PDAs almost 95% of the total battery power is needed for backlighting. This requires a very efficient backlight inverter, even more than in notebooks in order to increase the battery run-time. Cold cathode fluorescent lamps (CCFL) are used for backlighting of the LCD and satisfy the requirements on display performance, size and efficiency. A sinusoidal voltage across and current through the lamp is preferred to minimize EMI and

maximize the lamp efficiency. While the displays get thinner, there is a tendency to restrict the space reserved for the backlight inverter to a very limited volume. Both height and width of the inverter are critical parameters. The transformer is usually the largest and highest component on the board. To avoid high voltage breakdown, the coil former must be quite long to provide enough distance between beginning and end of the secondary winding. Also it should have multiple sections to keep the winding capacitance within reasonable limits. Optimization of the magnetics for maximum throughput power under these conditions leads to very long and narrow core designs in which traditional core shapes like E or EFD are not always suitable anymore. Any deformation or shifting of the long core legs will cause variations in inductance of the transformer due to misalignment of the narrow mating faces. This makes

it difficult to achieve a repeatable production of transformers or inductors. To avoid this problem, the magnetic circuit can be constructed with a closed ferrite frame and flat bar. In this way deformation does not play a major role anymore and mating faces have a larger surface area. Mounting of the transformer has become easy while inductance values are more repeatable. The frame and bar cores (including dedicated SMD coil formers) are standard available in 4 sizes and made of the low loss power ferrites 3C90 and 3C91. Other sizes are available on request. The 9 mm wide Frame cores can drive a lamp of approximately 3-4 W, the 10 mm wide core is suitable for 4-5 W lamps and the 12 mm wide type does 5-6 W. They are intended for use in Notebook PCs. In larger LCD monitors, 6-8 W is required for usually 2 or more backlighting lamps. The 15 mm wide core set is designed for this power level.



Fig.2 Example of Notebook computer with LCD backlighting

2. Design example

2.1 General

The backlight inverter described in this application note is based upon the principle of the Royer oscillator. As this oscillator is very basic with hardly no protections, nowadays most inverters make use of dedicated control ICs and a resonant tank to provide the high voltage transformer with a regulated sinusoidal voltage.

The schematic diagram is shown in figure 3. Monitoring the voltage across R_{cur} gives information about the lamp current but it also detects fault conditions like broken lamps. Another basic feature of the IC is a dim function. The basic operation of the backlight inverter is as follows: the DC input voltage is inverted into an alternating square wave voltage by the control IC and the two mosfets and applied to the resonant tank. The key component inside the tank is the transformer, which generates in combination with the ballast capacitor C_s and the lamp's parasitic capacitance C_p the sinusoidal ignition and burn voltage for the lamp. The ignition voltage is the minimum voltage required to ignite the CCFL. CCFLs require two to three times the operating voltage to ignite and this is generally dependent on length and thickness of the lamp. The exact ignition voltage is dependent on the age of the lamp and its ambient temperature. CCFL suppliers specify a worst case ignition voltage at the lamp's end of life and this value must be used in the design of the high voltage transformer.

In the next sections a complete design example of a backlight inverter for a 14" notebook

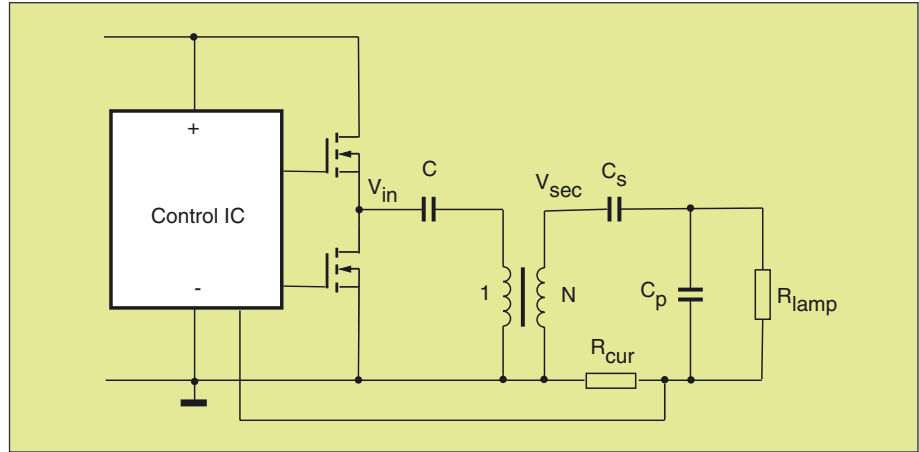


Fig.3 Schematic diagram of the backlight inverter

is worked out. This screen size requires a 220 mm long CCFL. The electrical properties of this CCFL are:

- Ignition voltage of 1400 Vrms
- Lamp voltage and current in burning state: 600 V and 5 mA (= 120 k Ω lamp resistance)

2.2. Choice of capacitance values on the secondary side

Each CCFL has parasitic capacitances to its surroundings. Nominal values for 14" lamps are 10..20 pF in burning state. In this example is chosen for 15 pF. The parasitic capacitance C_p needs to be determined for each lamp. As the parasitic capacitances become more dominant at higher frequencies, normally the operating frequency of most backlight inverters is between 50 and 150 kHz, while the burn frequency is close to 50 kHz. After a value has been determined for C_p , the value for the ballast capacitor C_s can also be chosen. A normal value for C_s to start with is 47 pF. As the optimization can easily be made by simulation, in practice some additional bench measurements have to be done to achieve a satisfied design.

2.3. Transformer design

2.3.1. Circuit analysis

A typical input voltage for backlight inverters for notebooks is 12V. The control IC in combination with the mosfets inverts this voltage into a square wave voltage between 12V and ground, present at the left side of DC-blocking capacitor C. Due to this capacitor, an alternating square wave voltage between +6V and -6V is present at the primary winding of the transformer. As capacitor C only has to block the DC-component of the input voltage, its value can be high and as a consequence the component can be ignored in the calculation of the transfer function of the resonant tank. Using the Fourier series, it can be proved that the first harmonic of the square wave can be written as:

$$V_1 = \frac{4}{\pi} V_{sq} \sin(2\pi ft) \quad [1]$$

in which V_{sq} represents the amplitude of the square wave voltage of 6V. The rms-value V_{in} of the sinusoidal voltage V_1 is equal to $1/\sqrt{2} \cdot V_1 = 5.4V$ and will be applied to the resonant tank. In order to calculate the inductance values and to analyze the resonant tank, first a simple model is introduced. The non-ideal transformer can be split up into a magnetizing inductance L_m , the leakage inductances at primary (L_{s1}) and secondary side (L_{s2}) and its turns ratio N, see figure 4.

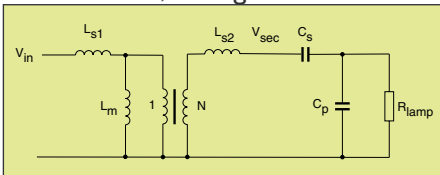


Fig. 4. Schematic diagram of the resonant tank

This diagram can further be simplified by transferring all primary components like the primary leakage inductance, the magnetizing

inductance, but also the input voltage V_{in} to the secondary side. This leads to the circuit shown in figure 5.

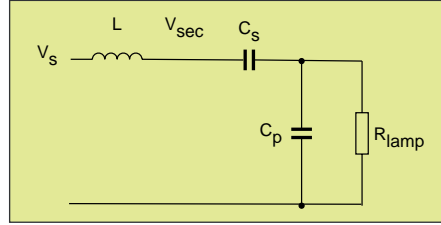


Fig. 5. Simplified schematic diagram

With this circuit the transfer function of the resonant tank will be determined. The input voltage V_{in} present at the primary side, transferred to the secondary side becomes:

$$V_s = \frac{L_m}{L_{s1} + L_m} N V_{in} = k N V_{in} \quad [2]$$

Voltage V_s is not the real secondary voltage, but the voltage present at the magnetizing inductance L_m transferred to the secondary side of the ideal transformer. The factor k, known as the couple factor, is a figure describing how much flux generated by the primary winding is caught up by the secondary. Or in other words: the magnetizing (or mutual) inductance L_m is equal to $L_{prim}k$, the primary leakage inductance is $L_{prim}(1-k)$ and the secondary leakage inductance is $L_{sec}(1-k)$. The ratio between the secondary inductance and primary inductance is equal to the square of the turns ratio: $N^2 = L_{sec}/L_{prim}$. As the primary and secondary windings are adjacent to each other in the coil former, the couple factor is normally between 0.4 and 0.7. In this design example, a couple factor of 0.6 has been chosen.

Inductance L, which is the series connection of L_{s2} and the parallel value of the primary inductances transferred to the secondary side, can be described as:

$$\begin{aligned} L &= L_{s2} + \frac{L_{s1} L_m}{L_{s1} + L_m} N^2 \\ &= L_{sec}(1-k) + \frac{L_{prim}^2 (k - k^2)}{L_{prim}} \cdot \frac{L_{sec}}{L_{prim}} \\ &= L_{sec}(1 - k^2) \end{aligned} \quad [3]$$

The series equivalent value C of the parasitic capacitance C_p and the ballast capacitor C_s is 11.37pF, see also section 2.2. Before ignition, the lamp resistance is several megohms and therefore figure 5 can even be more simplified to the schematic diagram of figure 6.

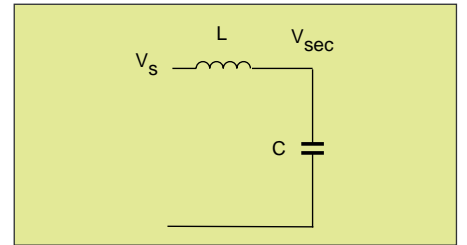


Figure 6. Simplified schematic diagram before ignition.

The resonant frequency f_0 of these reactive components at the secondary side is:

$$f_0 = \frac{1}{2\pi\sqrt{LC}} \quad [4]$$

Choosing a resonant frequency of 55kHz, the value for L can be calculated and is 736mH. The couple factor of 0.6 gives a secondary inductance L_{sec} of 1.15H, see formula [3]. Table 1 shows the components and parameters which have been determined so far:

C_s	47pF
C_p	15pF
k	0.6
V_{in}	5.4V
L	736mH
L_{sec}	1.15H
f_0	55kHz

Table 1. Determined circuit values.

2.3.2. Transfer function

The transfer function of the resonant tank in figure 5 can be described with:

$$H(j\omega) = \frac{V_{\text{lamp}}}{V_s} \quad [5]$$

$$= \frac{1}{1 - \omega^2 L C_p + \frac{C_p}{C_s} + \frac{j\omega L}{R_{\text{lamp}}} + \frac{1}{j\omega R_{\text{lamp}} C_s}}$$

A complete derivation of this transfer function, including the determination of V_s and operating frequencies can be found in appendix A. Resistance R_{lamp} follows from the lamp properties (600 V/5 mA) in burning state. As the lamp resistance before ignition is very high, the “not ignited” transfer function can be obtained by omitting the two imaginary parts in formula [5]. As V_s is 935 V (see appendix A) the relation between the lamp voltage and frequency can be made and is shown in figure 7

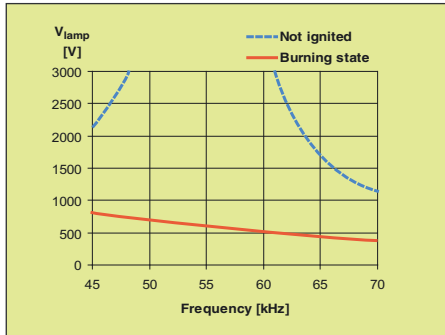


Fig. 7. Lamp voltage as function of frequency

With formula [2], the turns ratio N can be determined and is 289. Before ignition, the operating frequency of the inverter starts at a frequency higher than the resonant frequency of the tank and is decreasing until the voltage across

the lamp is sufficient to ignite the lamp.

The transfer function between the real secondary voltage V_{sec} and the lamp voltage (see fig. 5) can easily be derived and is equal to:

$$G(j\omega) = \frac{V_{\text{lamp}}}{V_{\text{sec}}} \quad [6]$$

$$= \frac{1}{1 + \frac{C_p}{C_s} + \frac{1}{j\omega R_{\text{lamp}} C_s}}$$

To ignite a 220 mm lamp for a 14” notebook, a worst case lamp voltage of about 1400 Vrms is needed. With equation [6], the secondary ignition voltage $V_{\text{sec-ign}}$ is 1847 V. The “not ignited” curve will reach a lamp voltage of 1400 V at a frequency of 67.5 kHz (as determined in appendix A), see figure 7.

With the ignition voltage, the secondary number of turns can be determined:

$$N_{\text{sec}} = \frac{V_{\text{sec-ign}} \sqrt{2}}{2 \pi f B_{\text{sat}} A_e} \quad [7]$$

The application temperature for backlight inverters in notebooks and PDAs is normally around 40 to 50 °C. For these temperatures 3C91 is the best material to use, because it has its minimum loss value around 60 °C.

Having a reasonable value for the maximum flux density in a 3C91 core of 330 mT (by assuming a worst case transformer temperature of 60 °C) and the minimum effective core area of the FRM27/3.8/9 core set of 8.7 mm², the formula returns 2145 turns for the secondary winding.

After ignition the lamp voltage drops to 600 V and the current stabilizes at 5 mA.

Most backlight inverters operate at the right side of the resonance peak in the so-called inductive area (the resonant tank is acting more inductive as capacitive) to obtain a safe switching behaviour for the mosfets.

The primary number of turns N_{prim} then is $N_{\text{sec}}/N = 2145/289 = 7.4$.

This figure is rounded to 7 and as a consequence the secondary number of turns reduces to $7 \times 289 = 2023$ and the maximum flux density during ignition increases up to 350 mT (formula [7]). With the secondary inductance L_{sec} and knowing that $N^2 = L_{\text{sec}}/L_{\text{prim}}$, the primary inductance is 13.8 μH.

Having this value, the core parameters of the FRM27/3.8/9 core set and the primary number of turns, the effective permeability of the core can be determined:

$$\mu_e = \frac{L_{\text{prim}} l_e}{\mu_0 N_{\text{prim}}^2 A_e} \quad [8]$$

in which μ_0 represents the permeability of free space ($= 4\pi \cdot 10^{-7}$ H/m).

This formula returns a value for μ_e of 1342. This leads to a small air gap, which is necessary to tune the primary inductance and keep its value within a specified tolerance.

2.4. Core losses

In burning state, the operating frequency of the backlight inverter is controlled by the IC and is 54 kHz (see appendix A).

According formula [6], the secondary voltage in burning condition is equal to 851 V. This leads to a flux density in burning state of 202 mT, which is significant lower than during ignition. The core volume V_e of the FRM27/3.8/9 core set is 504 mm³. (see appendix C) The core loss density can be calculated with the following fit formula:

$$P_{\text{core}} = c_m c_t f^x B^y \quad [9]$$

with frequency in Hz and flux density in T.

The fit parameters for 3C90 and 3C91 are shown in table 2:

	3C90	3C91
c_m	3.2×10^{-3}	3.5×10^{-3}
c_t	1	0.61
x	1.46	1.4
y	2.75	2.5

Table 2. Fit parameters for 3C90 and 3C91.

Using formula [9], the specified core loss density for 3C91 at a flux density of 202 mT, a frequency of 54 kHz and a transformer temperature inside the application of 60 °C is equal to 165 mW/cm³. From this, it follows that the core loss is 165 mW/cm³ × 0.504 cm³ = 83 mW.

L_{prim}	13.8 μH
L_{sec}	1.15 H
N_{prim}	7
N_{sec}	2023
N	289
μ_e	1342
l_e	52.1 mm
A_e	8.7 mm ²
V_e	504 mm ³

Table 3. Calculated values for the transformer properties.

	Ignition state	Burning state
V_{lamp}	1400 V	600 V
I_{lamp}	0	5 mA
R_{lamp}	∞	120 kΩ
V_{sec}	1847 V	851 V
V_s	935 V	935 V
f	67.5 kHz	54 kHz
B_{max}	350 mT	202 mT
P_{core}	450 mW	83 mW

Table 4. Calculated values for the ignition and burning state.

2.5. Winding design

In order to complete the design, one has to calculate the wire thickness for both primary and secondary side and the total copper loss. Essential data for the electrical design are: the winding area and the average turn length. The data for the FRM27/3.8/9 SMD bobbin are:

- Winding area A_w (primary): 1.75 mm² and A_w (secondary): 5 slots of 1.7 mm²,
- Average turn length l_{avg} : 18.5 mm

The total length l_{tot} of a wire can be calculated and is equal to the product of the number of turns of the primary or secondary winding and the average turn length l_{avg} . The maximum cross section A of the primary wire (including its insulation) is 1.75 mm² divided by 7 turns and is equal to 0.25 mm². From IEC wire tables, a wire having an overall cross section less than 0.25 mm² is chosen. In this case the wire diameter will be 0.45 mm and its copper cross section is 0.159 mm².

The secondary winding of 2023 turns is split up over the 5 sections and four sections will have 404 turns, while one slot will contain 407 turns. For the secondary side the cross section of the wire is 0.0042 mm², so a wire diameter of 0.050mm can be used with a copper cross section of 0.00196 mm².

The resistance of a winding can be determined by the following formula:

$$R_{\text{Cu}} = \rho \cdot \frac{n l_{\text{avg}}}{A} \quad [10]$$

in which ρ represents the resistivity of copper (20·10⁻⁹ Ωm at 60°C transformer temperature) and n

the number of turns of the primary or secondary winding. The copper resistances are shown in figure 8. The core losses, calculated in the previous section are represented by R_{core} .

2.6. Copper losses

In order to calculate the copper losses inside the transformer, first the currents through the resonant tank of figure 8 have to be determined. From figure 8 it is obvious that the input current of the transformer (flowing through L_{s1}) is not simply the secondary current multiplied by the turns ratio. Resistor R_{core} (representing the core losses), but especially magnetizing inductance L_m has a significant contribution to the input current. The input current is the scalar sum of the currents through L_m , R_{core} and the primary winding of the ideal transformer. In appendix B a detailed analysis of calculating the currents is given for both primary and secondary side. The copper loss

P_{Cu} can be calculated by [11]:

$$P_{Cu} = I^2 R_{Cu} \quad [11]$$

with I is the input current of 2.08 A or secondary current of 5.86 mA (both found in appendix B). The primary copper loss equals 70.5 mW. The total secondary loss is about 13 mW. The total copper losses then are about 84 mW.

2.7. Losses, efficiency and temperature rise of the transformer

The total losses of the transformer are the sum of the core- and copper losses and are less than 167 mW. As the total losses of the backlight inverter are also dependent on the capacitors, mosfets and control IC used, it is obvious that for a high efficiency of the complete design low loss components have to be used. With the lamp properties in burning state (600 V/ 5mA), the efficiency of the transformer becomes $3 W/3.167 W = 94.7 \%$.

The thermal resistance R_{th} which has been determined empirically for the wound Frame & Bar combinations can be described by:

$$R_{th} = \frac{1}{19 \cdot \sqrt{V_e}} \quad [12]$$

The effective volume (V_e) has to be inserted in units of cm^3 which returns R_{th} as $^{\circ}C/mW$. In thermal equilibrium the total losses in a transformer (P_{tot}) can be related to the temperature rise (ΔT) by the following relation:

$$P_{tot} = \frac{\Delta T}{R_{th}} \quad [13]$$

As the total losses are equal to 167 mW, the temperature rise of the transformer is $12.4^{\circ}C$.

Assuming an ambient temperature of $50^{\circ}C$, it will result in an operating temperature of $62^{\circ}C$ for the core set. If higher ambient temperatures are required, 3C90 (having minimum losses around $100^{\circ}C$) can be a good alternative.

Primary side		Secondary side	
Wire size	0.45 mm	Wire size	0.050 mm
Wires/slot	7	Wires/slot	404 x 4 + 407
R_{Cu1}	16.3 m Ω	R_{Cu2}	382 Ω

Table 5. Wire sizes, number of wires per slot and resistances.

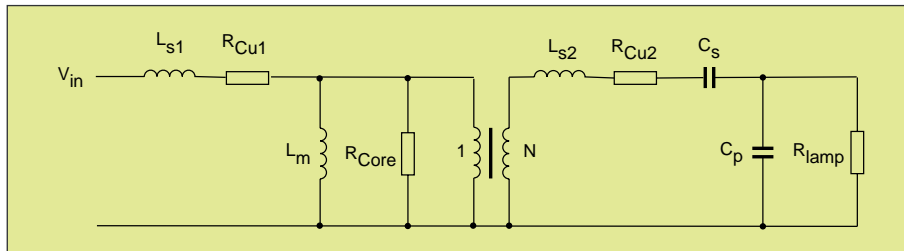


Fig. 8. Schematic diagram of the resonant tank including loss components.

P_{core}	83 mW
P_{Cu}	84 mW
P_{tot}	167 mW
η	94.7 %
ΔT	12.4 $^{\circ}C$

Table 6. Total losses, efficiency and temperature rise of the transformer.

Appendix A

Derivation of the transfer function of the resonant tank

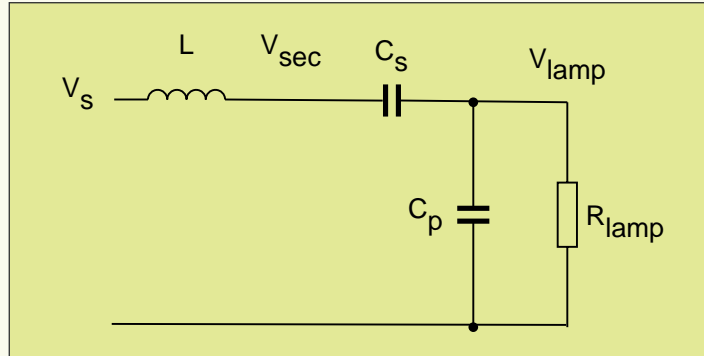


Fig. A.1. Schematic diagram of the resonant tank.

To determine the transfer function $H(j\omega) = V_{lamp} / V_s$ of the schematic diagram shown in figure A.1., first the impedance of the lamp (including its parasitic capacitance C_p) is derived:

$$Z_{lamp}(j\omega) = \frac{R_{lamp} \cdot \frac{1}{j\omega C_p}}{R_{lamp} + \frac{1}{j\omega C_p}} = \frac{R_{lamp}}{1 + j\omega R_{lamp} C_p} \quad [A.1]$$

By considering the series reactance of L and ballast capacitor C_s as $Z_s(j\omega)$:

$$Z_s(j\omega) = j\omega L + \frac{1}{j\omega C_s} \quad [A.2]$$

the transfer function $H(j\omega)$ can be written as:

$$H(j\omega) = \frac{Z_{lamp}(j\omega)}{Z_{lamp}(j\omega) + Z_s(j\omega)} = \frac{\frac{R_{lamp}}{1 + j\omega R_{lamp} C_p}}{\frac{R_{lamp}}{1 + j\omega R_{lamp} C_p} + j\omega L + \frac{1}{j\omega C_s}} \quad [A.3]$$

Multiplying both nominator and denominator with $(1 + j\omega R_{lamp} C_p)$ leads to:

$$H(j\omega) = \frac{R_{lamp}}{R_{lamp} + j\omega L - \omega^2 R_{lamp} L C_p + \frac{1}{j\omega C_s} + \frac{R_{lamp} C_p}{C_s}} \quad [A.4]$$

Dividing by R_{lamp} leads to the transfer function used in this application note:

$$H(j\omega) = \frac{V_{lamp}}{V_s} = \frac{1}{1 - \omega^2 L C_p + \frac{C_p}{C_s} + \frac{j\omega L}{R_{lamp}} + \frac{1}{j\omega R_{lamp} C_s}}$$

[A.5]

In this stage of the design example, voltage V_s (which is determined by kNV_{in} , see formula [2] in section 2.3.1.) has not been determined yet, because the turns ratio N is still unknown. But important to know is that V_s is frequency independent. The other parameters and component values of the tank are shown in table A.1.

Ignition state		Burning state	
V_{lamp}	1400 V	V_{lamp}	600 V
R_{lamp}	∞	R_{lamp}	120 k Ω
L	0.736 H	L	0.736 H
C_s	47 pF	C_s	47 pF
C_p	15 pF	C_p	15 pF

Table A.1. Lamp voltage and component values in the resonant tank.

In burning state V_s is equal to:

$$V_s = \frac{V_{lamp}}{H(j\omega)} = V_{lamp} \cdot \sqrt{\left(1 - \omega^2 L C_p + \frac{C_p}{C_s}\right)^2 + \left(\frac{\omega L}{R_{lamp}} - \frac{1}{\omega R_{lamp} C_s}\right)^2}$$

[A.6]

In ignition state, the lamp resistance is infinite, and voltage V_s can be written as:

$$V_s = \frac{V_{lamp}}{H(j\omega)} = V_{lamp} \cdot \left(1 - \omega^2 L C_p + \frac{C_p}{C_s}\right)$$

[A.7]

Normally the operating frequency in burning state is close to the minimum frequency of 50 kHz, see section 2.2. Choosing an operating frequency in burning state of 54 kHz, results with formula [A.6] in a voltage V_s of 935 V. With this voltage, the ignition frequency can be determined with formula [A.7] and equals 67.5 kHz.

V_s	935 V
f_{burn}	54 kHz
f_{ign}	67.5 kHz

Table A.2. Voltage V_s and frequencies in burning and ignition state.

Appendix B

Determination of currents and voltages in the resonant tank

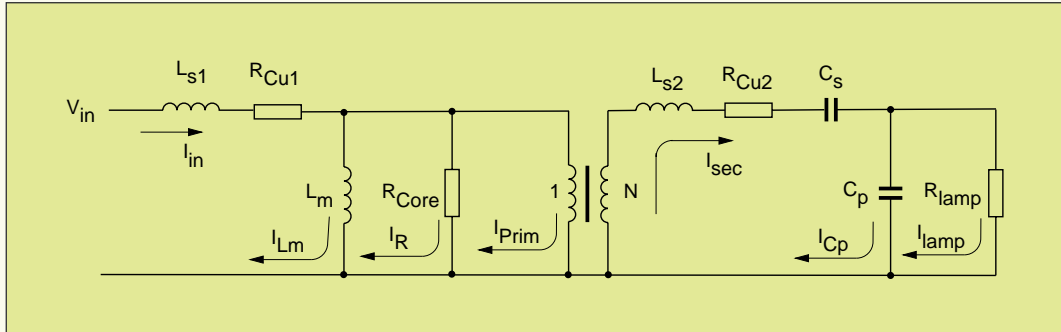


Figure B.1. Schematic diagram of the resonant tank including loss components.

With the use of table B.1, which gives the lamp resistance and the reactances of the other components at the burning frequency of 54 kHz, the currents in the resonant tank of figure B.1. will be determined. Compared to the values of the other secondary components, copper resistance R_{Cu2} ($= 382 \Omega$) is very small. The same holds for the reactance of the primary leakage inductance and R_{Cu1} ($= 16.3 \text{ m}\Omega$). Their contribution in calculating the currents is negligible and for this reason R_{Cu1} and R_{Cu2} will not be used in the determination of the primary and secondary current.

R_{lamp}	120 k Ω
X_{Cp}	196.5 k Ω
X_{Cs}	62.7 k Ω
$X_{Ls1} = X_{Lprim}(1-k)$	1.87 Ω
$X_{Ls2} = X_{Lsec}(1-k)$	156.1 k Ω

Table B.1. Resistance and reactances of the circuit at the burning frequency of 54 kHz.

As the lamp voltage is 600 V in burning state, the current I_{Cp} through the parasitic capacitance C_p can be calculated and is equal to $600 \text{ V} / 196.5 \text{ k}\Omega = 3.05 \text{ mA}$. The total secondary current I_{sec} delivered by the transformer is the scalar sum of this parasitic current and the current through R_{lamp} (5 mA) and equals 5.86 mA. Due to the influence of the capacitor C_p , current I_{sec} will lead on the lamp current I_{lamp} with a phase angle which is equal to $\arccos(5/5.86) = 31.4^\circ$. Putting real values on the horizontal axis and imaginary values on the vertical axis, the relation between voltages and currents can be displayed in a vector diagram, see figure B.2.

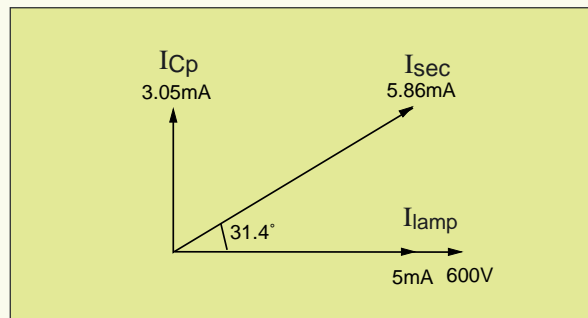


Fig. B.2. The relation between the lamp voltage and –current.

The secondary current is flowing through ballast capacitor C_s and secondary leakage inductance L_{s2} . As the reactance of the leakage inductance is greater than the reactance of C_s (see table B.1.), the equivalent reactance $X (= X_{L_{s2}} - X_{C_s})$ of both components is $93.4 \text{ k}\Omega$ and is inductive. The voltage across the equivalent reactance X can be calculated with $I_{\text{sec}} \cdot X$ and is 547 V , see also figure B.3. Because of the inductive behaviour of X , the voltage is 90° ahead on the secondary current. The voltage can be split up into a real and an imaginary part, see figure B.3. Due to the 90° phase angle between voltage X and the secondary current, the phase angle of 31.4° can also be found between the voltage of 547 V and the imaginary part of this voltage. The imaginary and real part can be calculated by $547 \text{ V} \cdot \cos(31.4) = 467 \text{ V}$ and $547 \text{ V} \cdot \sin(31.4) = 285 \text{ V}$. The real part of the voltage across the series connection of X and the lamp impedance is $600 \text{ V} - 285 \text{ V} = 315 \text{ V}$. The secondary voltage $V_{\text{sec-ideal}}$ of the ideal transformer is then $\sqrt{(315 \text{ V})^2 + (467 \text{ V})^2} = 563 \text{ V}$. The phase angle between voltage $V_{\text{sec-ideal}}$ of 563 V and the 600 V lamp voltage is $\arccos(315/563) = 56^\circ$. And as a result the phase angle between the voltage of 563 V and the secondary current is $56^\circ - 31.4^\circ = 24.6^\circ$

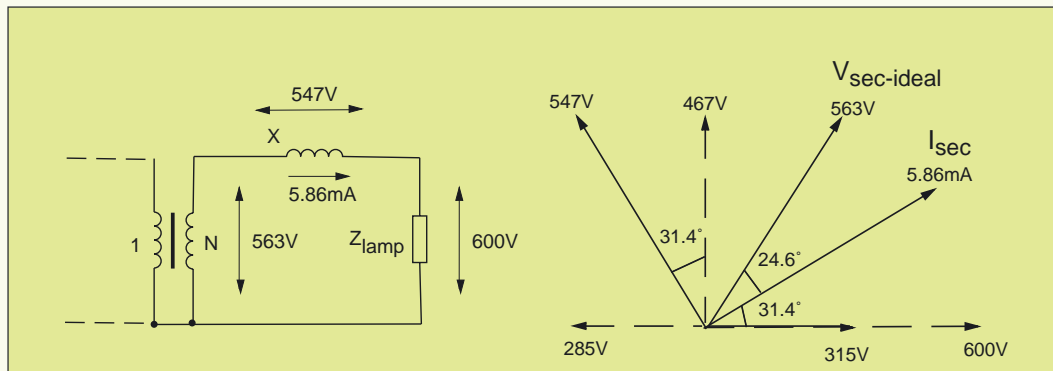


Fig. B.3. The relation between secondary voltages and -current.

Voltage V_{L_m} on the primary side can be found by dividing $V_{\text{sec-ideal}}$ by the turns ratio N of 289 and equals 1.95 V . This voltage is put on the horizontal axis, because it is present across the "core loss" resistor R_{core} , see figure B.4. The current through the primary of the "real" transformer ($I_{\text{prim-ideal}}$) can be found by multiplying the secondary current of 5.86 mA with the turns ratio and equals 1.7 A . The phase angle between voltage V_{L_m} and current $I_{\text{prim-ideal}}$ is still 24.6° , see figure B.4. Resistor R_{core} , representing the core losses (table 3), is connected in parallel with the magnetizing inductance L_m . The current I_R through R_{core} is $83 \text{ mW} / 1.95 \text{ V} = 43 \text{ mA}$. Current $I_{\text{prim-ideal}}$ of 1.7 A can be split up into a real part of 1.55 A and an imaginary part of 0.71 A , see figure B.4. Knowing that $B = \mu H = \mu N I / l_e$, the current through L_m can be determined. With the effective flux density B of $202 \text{ mT} / \sqrt{2} = 143 \text{ mT}$ in burning state, the current can be determined (see also table 3):

$$I_{L_m} = \frac{B l_e}{\mu_0 \mu_e N_{\text{prim}}} \quad \text{[B.1]}$$

and has a value of 0.63 A .

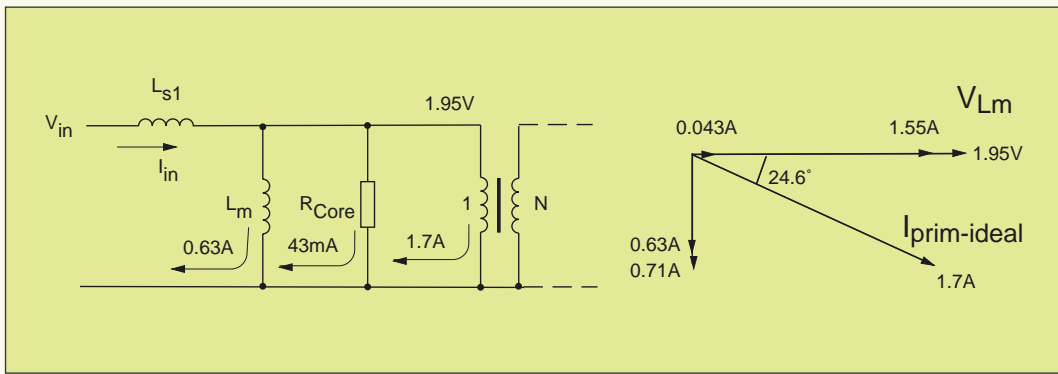


Fig. B.4. Voltages and currents at the primary side.

The current through the primary leakage inductance, and thus the current I_{in} flowing into the transformer, is the scalar sum of the real part ($43 \text{ mA} + 1.55 \text{ A} = 1.59 \text{ A}$) and the imaginary part ($0.71 \text{ A} + 0.63 \text{ A} = 1.34 \text{ A}$) in figure B.4 and equals 2.08 A . The phase angle between the real part of 1.59 A and the input current of 2.08 A is $\arccos(1.59/2.08) = 40.1^\circ$, see also figure B.5.

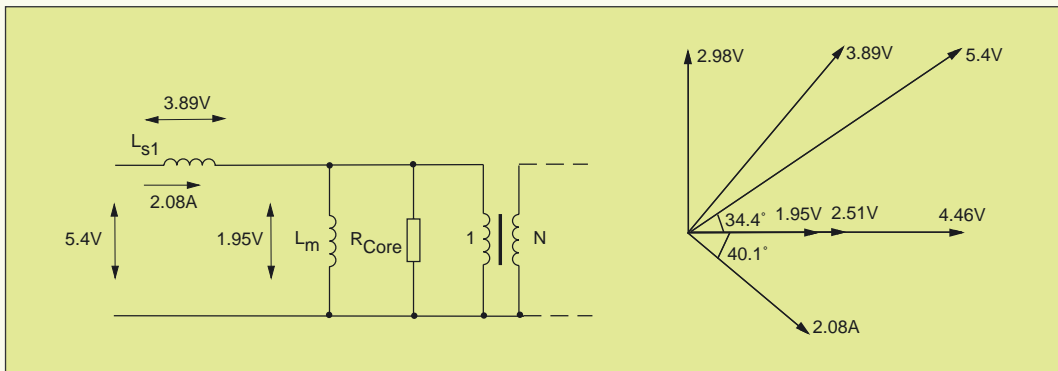


Fig. B.5. Input voltage and -current.

The voltage across the primary leakage inductance L_{s1} is equal to $2.08 \text{ A} \cdot 1.87 \Omega = 3.89 \text{ V}$. This voltage is 90° ahead on the input current, the angle between this voltage and its real part is $90^\circ - 40.1^\circ = 49.9^\circ$. Again the voltages of the real and imaginary part can be calculated and are 2.51 V and 2.98 V respectively. The total real part of the input voltage V_{in} is the sum of 2.51 V and 1.95 V and is 4.46 V .

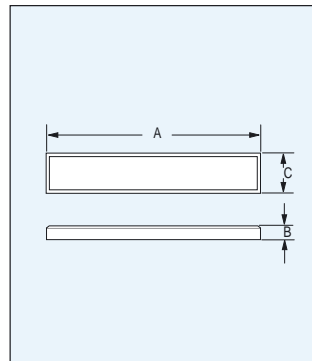
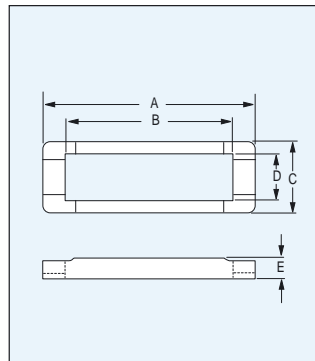
Having the imaginary part of 2.98 V , input voltage V_{in} is equal to $\sqrt{(2.98 \text{ V})^2 + (4.46 \text{ V})^2} = 5.4 \text{ V}$. This is a known voltage, see section 2.3.1., because it was a parameter to start the design with. The phase angle between the input voltage and its real part is $\arccos(4.46/5.4) = 34.4^\circ$ and the angle between the input voltage and current is $34.4^\circ + 40.1^\circ = 74.5^\circ$.

In the table below, the currents necessary to calculate the copper losses can be found.

I_{sec}	5.86 mA
I_{in}	2.08 A

Table B.2. Secondary and input current of the resonant tank.

Survey of the available Frame & Bar core range



Cores with a design similar to Frame and Bar cores have been available from Philips under the name of H cores, since 1971. They were mainly applied as signal transformers in Telecom applications. The new Frame and Bar cores have been modified to a slim and elongated rectangular shape in order to meet the dimension requirements of a flat LCD panel. The elongated rectangular shape is also optimized to accommodate the large number of turns required to generate the high ignition voltage (1400 Vrms) for a backlight discharge lamp. Besides this, the Frame and Bar core is also easy to assemble into a transformer and has been adopted as a standard core for the LCD backlight inverter transformer. A backlight inverter is an electronic DC to AC circuit that drives a Cold Cathode Fluorescent Lamp (CCFL) for the backlighting of a notebook LCD display or LCD monitor .

Summary :

- ◆ Narrow design
- ◆ Easy to assemble
- ◆ Large winding space to accommodate a high number of turns

Core type		FRM 20/5/15	FRM 21/4/12	FRM 24/3.9/10	FRM 27/3.8/9	BAR 20/3/5.5	BAR 22/2/6	BAR 25/2.2/4	BAR 28/3.8/2.3
effective core parameters	core factor $\Sigma I/A(\text{mm}^{-1})$	3.29	5.06	5.65	5.56	3.29	5.06	5.65	5.56
	eff. volume $V_e (\text{mm}^3)$	655	312	370	504	655	312	370	504
	eff. length $l_e (\text{mm})$	46	40	45.8	52.1	46	40	45.8	50
	eff. area $A_e (\text{mm}^2)$	14	7.9	8.1	9.7	14	7.9	8.1	9.0
	min. area $A_{\text{min}} (\text{mm}^2)$	7.4	5.7	6	8.7	7.4	5.7	6	8.7
	mass of core half (g)	≈ 2.1	≈ 1.5	≈ 1.3	≈ 1.6	≈ 1.5	≈ 1	≈ 1.2	≈ 1.2
dimensions (mm)	A	19.7 ± 0.3	21 ± 0.2	23.8 ± 0.3	26.7 ± 0.7	19.9 ± 0.3	21.8 ± 0.3	24.7 ± 0.3	28 ± 0.5
	B	15.6 ± 0.3	16.2 ± 0.3	19.2 ± 0.3	19.7 ± 0.6	2.85 ± 0.05	1.8 ± 0.1	2.15 ± 0.05	3.8 ± 0.1
	C	14.8 ± 0.3	11.8 ± 0.25	9.8 ± 0.2	9.0 ± 0.3	5.45 ± 0.15	5.5 ± 0.2	4.4 ± 0.2	2.3 ± 0.1
	D	11.4 ± 0.25	8.9 ± 0.2	7.3 ± 0.2	6.5 ± 0.2				
	E	4.6 ± 0.1	4.0 ± 0.1	3.85 ± 0.1	3.8 ± 0.2				
coil formers	CPHS	■	■	■	■				
mounting parts	COV	■							

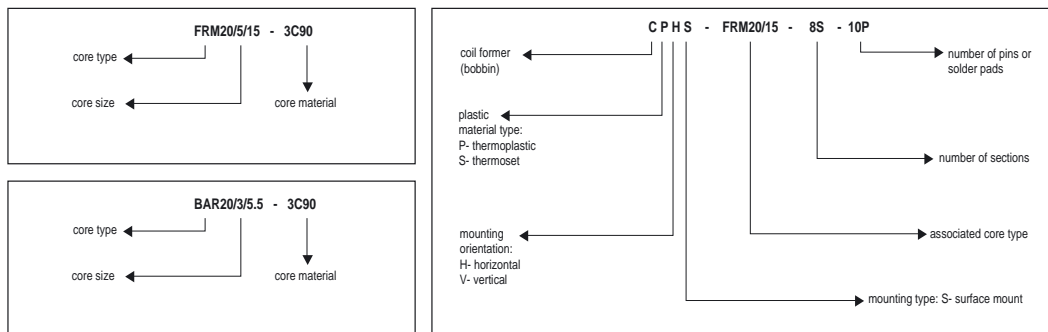
For full details of cores and bobbins see our Data Handbook or visit our web site at www.ferroxcube.com

Core type	FRM20/5/15	FRM21/4/12	FRM24/3.9/10	FRM27/3.8/9	
Matching cores	BAR20/3/5.5	BAR22/2/6	BAR25/2.2/4	BAR28/3.8/2.3	
core SETS for power applications	3C90	500	400	370	350
	3C91 prot	600	470	440	420

500 — ungapped core set. $A_L = 500$

A_L value (nH) measured at $\dot{B} \leq 0.1$ mT, $f \leq 10$ kHz, $T = 25^\circ\text{C}$

A_L tolerance: $\pm 25\%$



Material characteristics

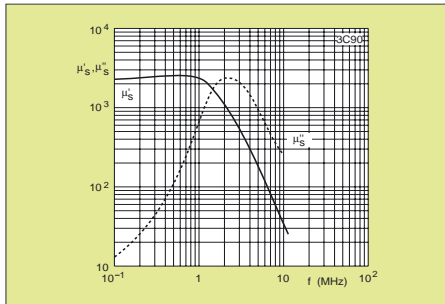
3C90 SPECIFICATIONS

SYMBOL	CONDITIONS	VALUE	UNIT
μ_i	25 °C; ≤ 10 kHz; 0.1 mT	$2300 \pm 20\%$	
μ_a	100 °C; 25 kHz; 200 mT	$5500 \pm 25\%$	
B	25 °C; 10 kHz; 250 A/ m	≥ 430	mT
	100 °C; 10 kHz; 250 A/ m	≥ 340	mT
P_V	100 °C; 25 kHz; 200 mT	≤ 80	kW/m ³
	100 °C; 100 kHz; 100 mT	≤ 80	
	100 °C; 100 kHz; 200 mT	≈ 450	
ρ	DC, 25 °C	≈ 5	Ωm
T_C		≥ 220	°C
density		≈ 4800	kg/m ³

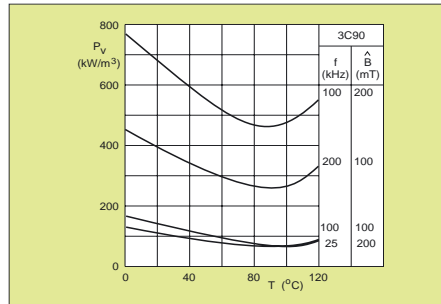
Properties measured on sintered, non ground ring cores of dimensions $\varnothing 25 \times \varnothing 15 \times 10$ mm which are not subjected to external stresses.

3C91 SPECIFICATIONS

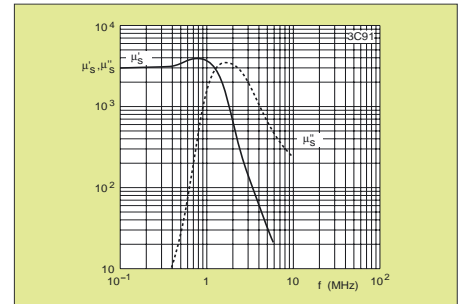
SYMBOL	CONDITIONS	VALUE	UNIT
μ_i	25 °C; ≤ 10 kHz; 0.1 mT	$3000 \pm 20\%$	
μ_a	100 °C; 25 kHz; 200 mT	$5500 \pm 25\%$	
B	25 °C; 10 kHz; 250 A/ m	≥ 430	mT
	100 °C; 10 kHz; 250 A/ m	≥ 330	mT
P_V	60 °C; 100 kHz; 100 mT	≤ 80	kW/m ³
	60 °C; 100 kHz; 200 mT	≤ 80	
ρ	DC, 25 °C	≈ 5	Ωm
T_C		≥ 220	°C
density		≈ 4800	kg/m ³



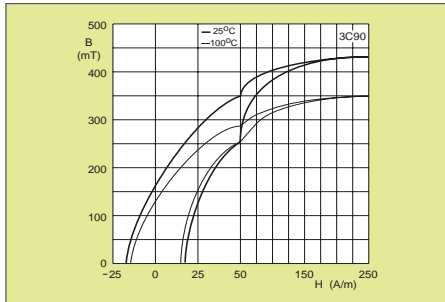
Complex permeability as a function of frequency



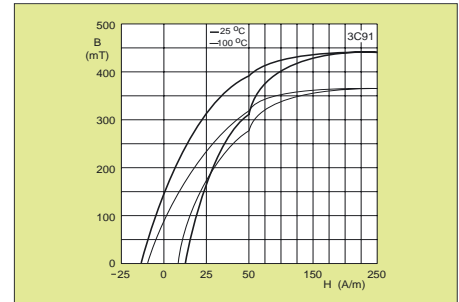
Specific power loss for several frequency/flux density combinations as a function of temperature.



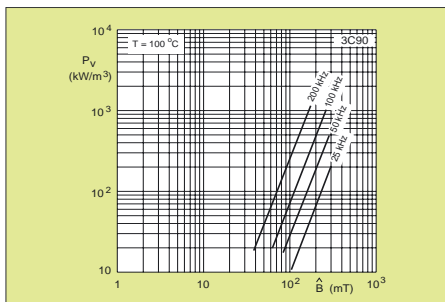
Complex permeability as a function of frequency



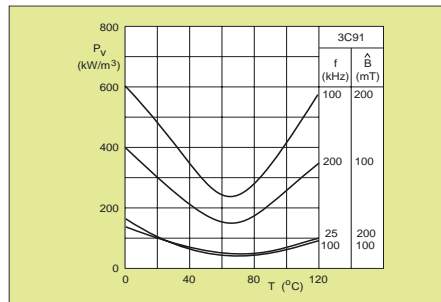
Typical B-H loops



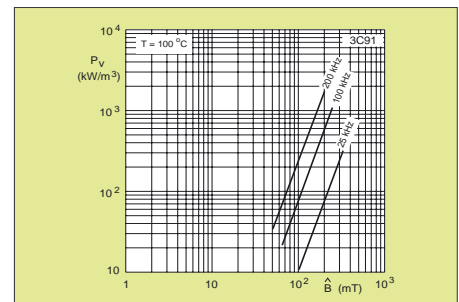
Typical B-H loops



Specific power loss as a function of peak flux density with frequency as a parameter.



Specific power loss for several frequency/flux density combinations as a function of temperature.



Specific power loss as a function of peak flux density with frequency as a parameter.