AN10971

TL applications with NXP ballast controllers

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Document information

Info	Content
Keywords	Half-bridge, electronic ballast, tube light
Abstract	This application note describes how to design an application to drive a fluorescent lamp with a half-bridge circuit combined with an NXP Semiconductors ballast controller.

Revision history

Rev	Date	Description
v.1	20110915	first draft

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1. Introduction

This application note describes the use of an NXP Semiconductors half-bridge driver IC for High Frequency (HF) Tube Light (TL) single tube, fixed DC bus voltage applications.

2. Scope

This application note is organized as follows:

- Section 3 describes the basic operation of a half-bridge ballast for fluorescent tubes.
- Section 4 describes how to select MOSFETs for the half-bridge.
- Section 5 describes how to select a resonant coil and capacitor.
- Section 6 describes how to design magnetic components.
- Section 7 describes the feedback control loop used in the dimmable controllers.
- Section 8 describes system performance.

3. Lamp characteristics and half-bridge principles

3.1 Introduction

Electronic ballasts must preheat, ignite, control and monitor the condition of the TL according to the TL specification. In cases of no ignition or lamp end of life, the ballast must shut down to avoid damage to the following:

- the ballast
- or the overheating of the lamp electrodes

Depending on the lamp type, tube ignition occurs at different voltage levels. For example, a Compact Fluorescent Lamp (CFL) ignition occurs at around 500 V and up to 1200 V for a TL. Following ignition a lamp is in its operating state. During operation lamp voltage is dependent on the shape and content of the tube. For TL lamps, voltages range from 80 V to 220 V (RMS).

A fluorescent lamp has a so called "negative incremental impedance". The more current flowing through a fluorescent lamp the lower the voltage (unlike a resistor). Driving a fluorescent lamp with a Constant Voltage (CV) would result in an unstable system. In addition, a breakdown of the supply and/or tube within a few milliseconds also occurs.

To stabilize the current through the lamp, a series impedance is needed typically a resistor. However, in practice a coil is used to reduce losses as shown in Figure 1. In magnetic 50 Hz to 60 Hz Low Frequency (LF) systems, see Figure 2 coils are large. However, in HF TL 40 kHz to 80 kHz systems coils are smaller as the impedance of a coil increases linearly with the frequency.

In steady state, at frequencies higher than 10 kHz a fluorescent lamp can be considered to be a resistor. At high frequencies Ohms law $V = I \times R$ is applicable to fluorescent tubes. The resistance (R) varies with the power supplied to the lamp. For example, a 36 W T8 lamp with an operating voltage of 100 V can be considered as a 277 Ω resistor. This condition is true when operated at its nominal power.

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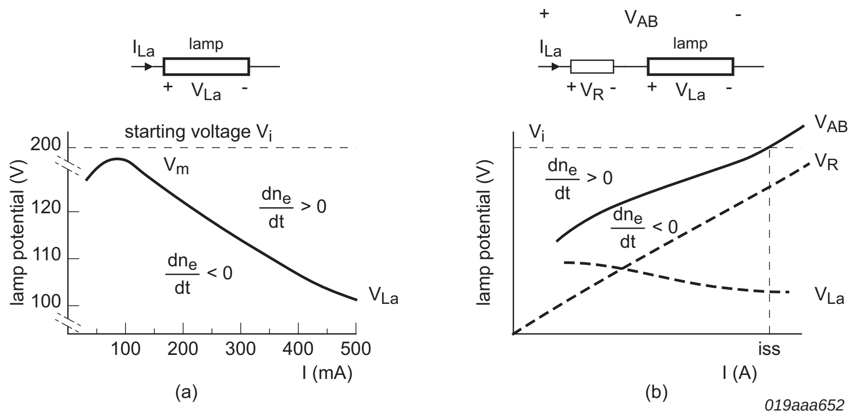

a. Discharge potential drop versus current

b. Effect of series resistance in stabilizing lamp current

(1) dn_e = delta in the number of free electrons in th plasma

Fig 1. Lamp characteristics

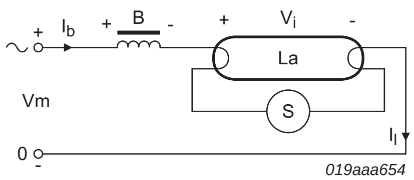

Fig 2. Magnetic 50/60 Hz TL ballast

3.2 Lamp filaments

Each lamp has two filaments (or electrodes), each consisting of a coated tungsten wire as shown in Figure 3. The resistance of the tungsten wire is directly related to its temperature. Figure 4 identifies the relationship between electrode temperature and the ratio of Resistance hot (R_h) and Resistance cold (R_c) of a typical tungsten wire.

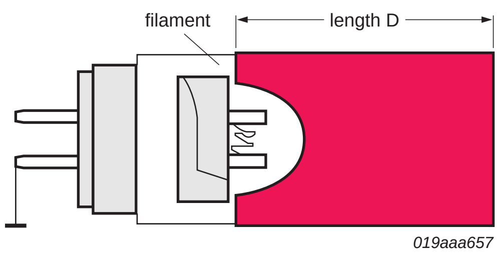

Fig 3. Lamp filament

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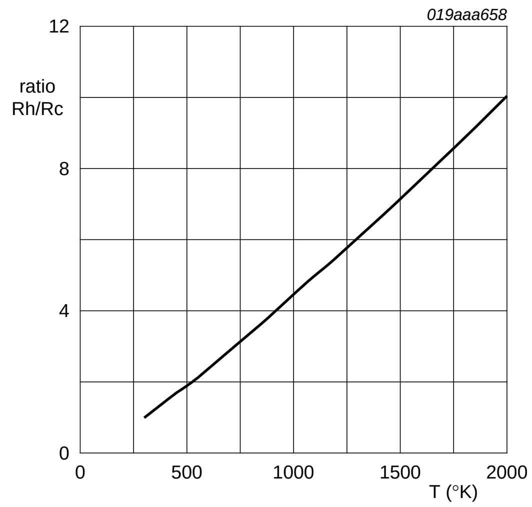

Fig 4. R_h/R_c ratio versus temperature for 100 % wolfram filaments

Lamp electrodes are preheated before the lamp is ignited or lamp life time is considerably reduced. During preheat, electrode resistance increases typically by a factor of between four to six. To generate sufficient and an even spread of heat before ignition, preheat time must not be lower than 0.5 s. Preheat times longer than 1.5 s to 1.7 s are considered undesirable as heat is lost to the gasses in the lamp tube.

Figure 5 identifies the operating area of fixed preheat current for a T8 36 W burner with two current levels. Figure 5 also shows that if the current is too high the optimal ignition temperature is reached before 0.5 s. As a result, either ignition takes place too early or the filament is overheated before ignition.

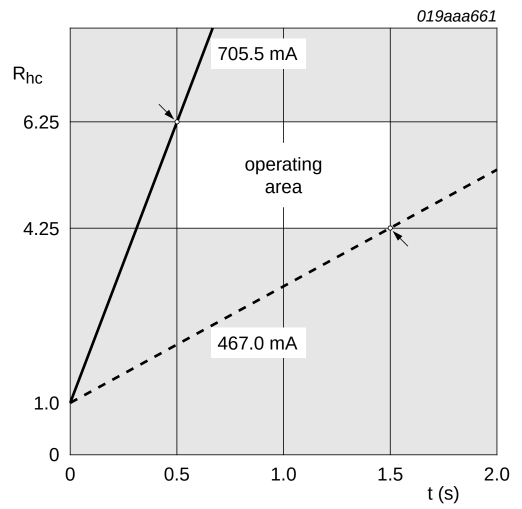

Fig 5. Fixed preheat current window

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3.3 HF TL half-bridge principles

High frequency TL applications typically run at 40 kHz to 60 kHz instead of 50 Hz to 60 Hz for magnetic ballast applications. In HF TL applications, the impedance of the coil is linear with its frequency. The size of a coil for an HF TL application is only a fraction of the coil required for 50 Hz to 60 Hz magnetic ballast applications.

Figure 6 (left) shows a theoretical 50 kHz ballast circuit with a 50 kHz sine wave generator. Figure 6 (right) shows a real 50 kHz ballast circuit with a DC blocking capacitor and square wave generator.

A 50 kHz sine wave is not easy to produce, therefore a 50 kHz square wave generator is used as shown in Figure 6 (right). The higher harmonics of the square wave hardly pass the coil whereas the lower harmonics, specifically the first harmonic does pass the coil. The square wave generator in combination with the coil in practice acts as a stable sine wave source. A square wave of 50 % duty factor and an amplitude between 0 V and V_{bridge} has a DC component of V_{bridge} / 2. This condition causes an unlimited DC current passing through the lamp. A DC blocking capacitor is used to filter out the DC voltage component to stop the DC current. The voltage on the DC blocking capacitor is V_{bridge} / 2.

The resonance frequency of the LC combination is used to ignite the lamp. Following ignition, the amplitude and frequency of the square wave together with the L and C values determine the lamp power.

The RMS value of the first harmonic of a square wave is:

$$V_{bridge} \times \left(\frac{\sqrt{2}}{\pi}\right)$$

Therefore, a square wave of 400 V corresponds with a sine wave of approximately 180 V (RMS). The higher harmonics (3,5,7) only contribute < 5 % to lamp current and can be ignored.

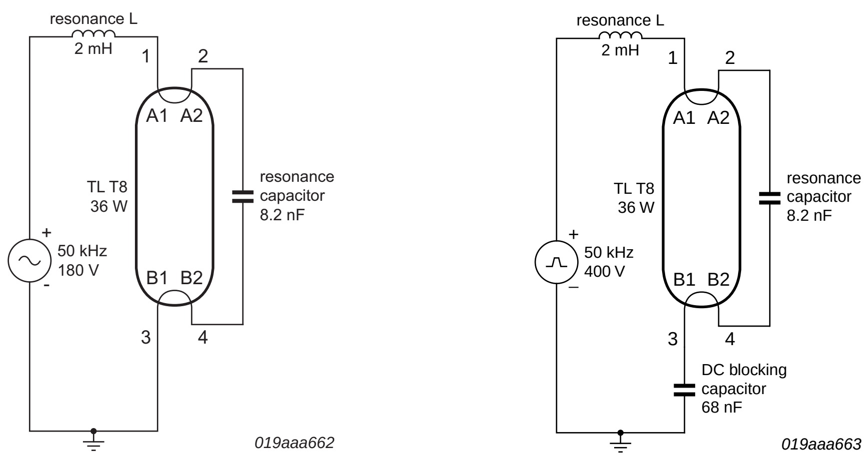

Fig 6. Half-bridge circuit overview

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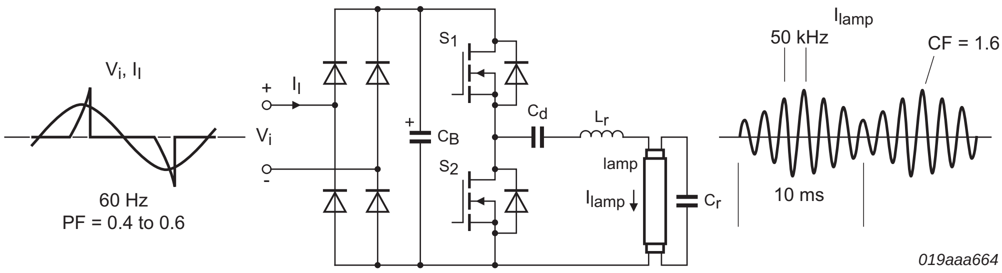
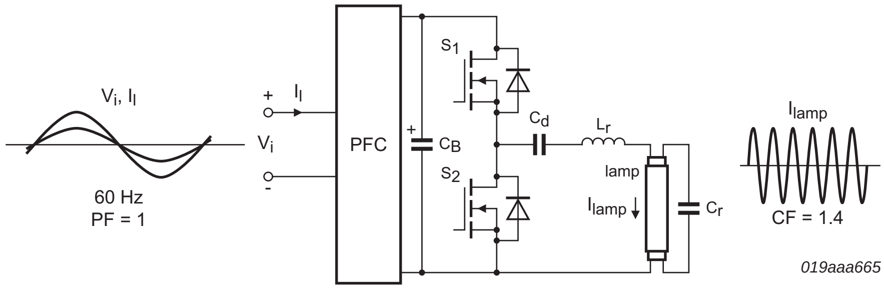


a. Typical HF TL half-bridge ballast using rectified mains



b. Typical HF TL half-bridge ballast using a PFC

Fig 7. Typical HF TL half-bridge ballast circuits

Figure 7 represents a typical HF TL half-bridge ballast circuit, so called as only two switches (S1 and S2) are used. Full-bridge designs, use four switches. Coil impedance, L_r as shown in Figure 7 limits the current in the circuit and can be calculated using Equation 1:

$$Z = 2 \times \pi \times frequency \times L_{coil} \tag{1}$$

As the impedance is linear to the frequency, a 1000 times higher frequency requires a 1000 times smaller coil value for the same impedance. As a result, a high frequency ballast is much smaller and lighter than a 50 Hz to 60 Hz magnetic ballast.

Switches S1 and S2 generate a square wave in the 40 kHz to 100 kHz range. The average voltage of a square wave is equal to half its amplitude. Capacitor (C_d) is a DC blocking capacitor, without it a DC current would flow through the coil and lamp. C_r and L_r together have a resonant frequency where the voltage sweeps up to a point where the lamp can be ignited. Capacitor C_r is placed between the filaments to preheat the filaments before ignition.

For more detailed information on how to select suitable resonant tank LC values see Section 5.

The supply feeding the half-bridge can be either rectified mains or a DC voltage generated by a pre-conditioner circuit. This type of circuits pulls current from the mains resembling the resistive load (so called Power Factor) of an incandescent bulb. In most countries higher than 25 W, a mains voltage-current source relation is mandatory.

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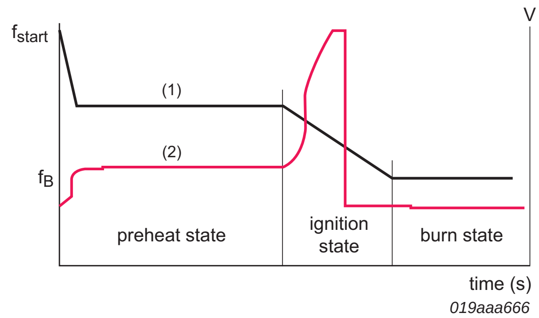
Power Factor Correction (PFC) circuits are either active (controller) or passive. The main passive PFC topologies used include, valley fill and charge pump. This application note assumes either rectified mains as shown in Figure 7 [A] or an active PFC as shown in Figure 7 [B].

If the square wave is a fixed frequency, and the voltage on the electrolytic capacitor fluctuates as in Figure 7 [A] the lamp current also fluctuates. The so-called crest factor of the lamp current is the peak/RMS value. To maximize the lifetime of the lamp, the crest factor must remain lower than 1.7.

3.4 Start-up, ignition and operation sequence



(1) Frequency versus time

(2) Lamp voltage versus time

Fig 8. Start-up sequence

Figure 8 shows an example of the start-up sequence for the UBA2021 controller which is similar to the start-up sequence of other NXP Semiconductors controllers.

At the start of the sequence, the controller starts at its maximum frequency (f_{start}) typically 100 kHz. The frequency then drops quickly to the preheat frequency of around 70 kHz to 80 kHz. The controller then waits for the preheat time to pass (typically 0.8 s to 1.5 s) then drops to a frequency of around 40 kHz (f_{min}).

The ignition (LC tank resonance) frequency of between 45 kHz to 55 kHz must occur between the end of preheat period and the minimum frequency of f_{min}. See Figure 8. Following ignition, that is, during the burn state the frequency of the half-bridge determines the power of the lamp. Figure 8 line (2) shows the voltage over the lamp.

3.5 Zero voltage switching of a half-bridge (capacitive mode)

MOSFET switches S1 and S2 need a precisely timed control voltage. The NXP Semiconductors controller generates this control voltage as shown in Figure 9.

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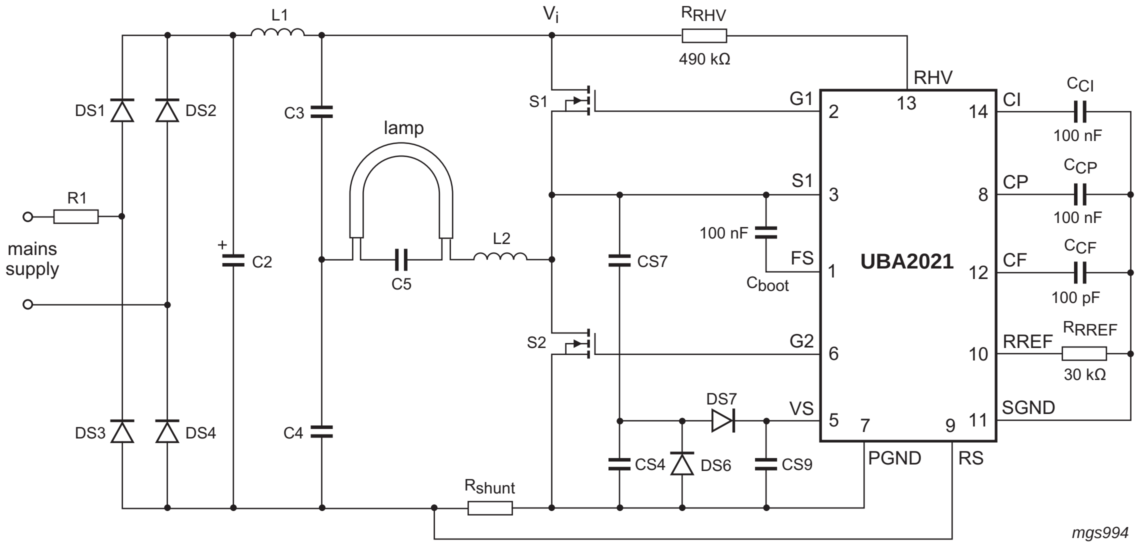

Fig 9. S1 and S2 driven by UBA2021

When the lamp is ignited, the phase shift of the Inductor Resistor Capacitor (LRC) L2, C5 and lamp combination above its resonance frequency is such that the current lags behind. The voltage current phase shift of an inductor is similar. Figure 10 shows the coil current amplitude of LRC resonant tank versus frequency while Figure 11 shows the phase relationship.

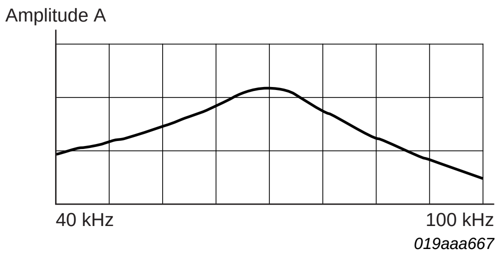

Fig 10. Current amplitude of LRC resonant tank versus frequency

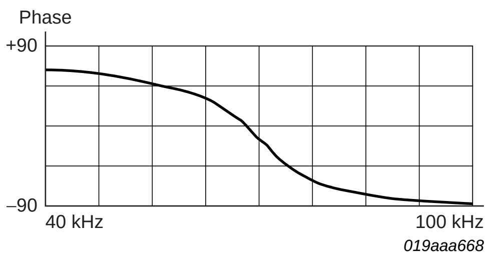

Fig 11. Current phase of LRC resonant tank versus frequency

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The behavior of the load is called inductive mode when the current lags behind on the voltage. For capacitive mode, the voltage lags behind on the current. These conditions are similar to the behavior of a capacitor and inductor. Figure 12 and Figure 13 show the phase relationship of voltage and current flowing through an inductor and a capacitor.

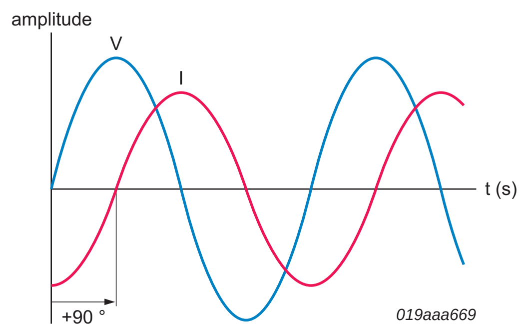

Fig 12. Current voltage relation of inductor

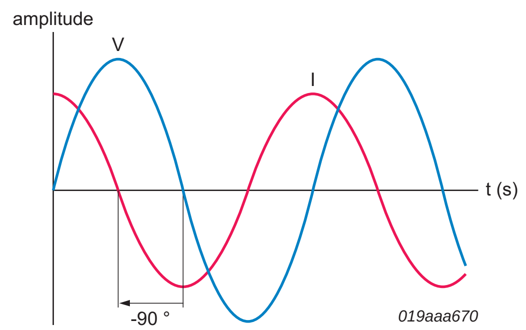

Fig 13. Current voltage relation of capacitor

If the current lags behind the voltage (inductive mode), the current charges and discharges the parasitic capacitance of the half-bridge when both MOSFETs are off. This state is known as the non-overlap time. The inductor current charges the parasitic capacitance for the low to high transition. The current flows through the bulk diodes of the high-side MOSFET until the MOSFET is switched on. For high to low transitions the opposite is true.

The result is that there are no switching losses in the MOSFETs since the MOSFETs are opened/closed when there is no voltage over them. This state is known as Zero Voltage Switching (ZVS), it is the way a half-bridge resonant tank normally operate.

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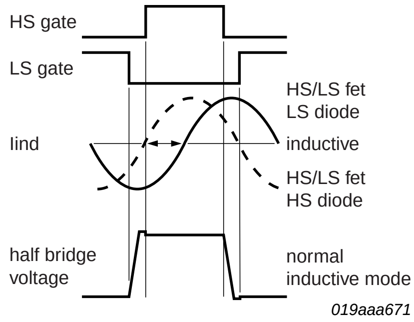

Fig 14. Switch timing sequence (inductive mode)

If the resonant tank is run lower than its resonant frequency, then it is in capacitive mode. In capacitive mode, the MOSFETs remain closed while large voltages exist across them. This results in large current spikes in the order of 10 A to 15 A. These spikes cause lower efficiency and can damage the MOSFETs. This condition is commonly known as "hard switching".

If the bulk diode in one MOSFET is conducting, current while the other MOSFET is opened the MOSFETs can be damaged within a single stroke. In capacitive mode, there is always full bus voltage hard switching. However, there can also be an in-between state with limited hard switching (of a voltage less than the bus voltage). All NXP Semiconductors controllers react to capacitive mode, albeit in a different way.

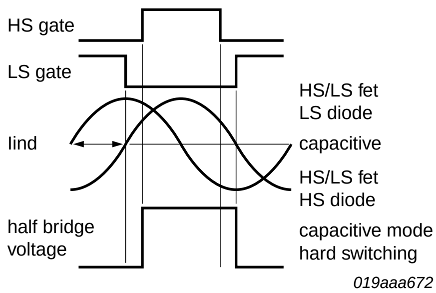

Fig 15. Switch timing sequence (capacitive mode)

The UBA2014 reacts to capacitive mode by jumping to its maximum frequency. The UBA2021 reacts by quickly increasing the frequency until there is no more hard switching. The UBA2015(A) and UBA2016A react by shutting down as capacitive mode is detected.

3.5.1 None overlap time, fixed versus adaptive

The time period when both MOSFETs are open (conducting) is called the non-overlap time. The non-overlap time is much larger than necessary to avoid cross conduction, when for example both MOSFETs are switched on. The time is set such that the resonant tank coil current has charged/discharged the capacitance on the half-bridge during the none overlap time. All extra current after this none overlap time, flow through the bulk diode of the MOSFET that is switched on after the none overlap time ends.

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The UBA2014 has an input that senses when the slope of the half-bridge square wave voltage has ended. The UBA2014 is able to adapt its timing such that it switches on the MOSFETs after the slope has ended. This condition minimizes the current through the bulk diodes of the MOSFETs and reduces losses. This principle is called "adaptive none overlap".

The UBA2015(A), UBA2016A and UBA2021 all have a fixed none overlap time. For a fixed non-overlap, the charge pump supply capacitor (CS7 in Figure 9) must be fine-tuned to eliminate any hard switching.

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4. Half-bridge MOSFET selection

4.1 Introduction

The half-bridge MOSFETs are both N type switched on/off at precisely the correct time. The current with which the gates are charged and discharged is also important for ElectroMagnetic Compatibility (EMC). The main parameters effecting MOSFET selection for use with half-bridge applications include:

- Breakdown voltage
- R_{on} resistance - On Resistance
- Low gate charge

The breakdown voltage of a MOSFET is a critical parameter, therefore selection must be within precise design limits. The R_{on} resistance value can be selected within limits depending on the target efficiency of the ballast. The low gate charge keeps the capacitor on the dV/dt supply small.

4.1.1 MOSFET voltage rating

Voltage overshoots in a half-bridge resonant tank always exist due to parasitics. Therefore, select a voltage rating of the MOSFETs that is 50 V higher than the bus voltage of the half-bridge.

4.1.2 MOSFET RON

The RMS current through the MOSFETs and the desired efficiency of the ballast determines the R_{on} of the MOSFET. R_{on} resistance depends on its operating temperature, for TL a typical value is 90 °C and 105 °C for CFL. The RMS current through the MOSFETs is the vector sum of the burner and resonant capacitor current.

As the lamp current is known that is, controlled or determined by a fixed frequency the resonant capacitor current is equal to Equation 2.

$$2 \times \pi \times f_{run} \times V_{lamp} \tag{2}$$

As the lamp and capacitor current have a 90° phase shift, the coil current is equal to Equation 3.

$$I_{coil} = \sqrt{(I_{lamp^2} + I_{capacitor^2})} \tag{3}$$

Losses in each of the MOSFETs do occur, the losses can be determined using Equation 4. Typical temperature values inside a TL ballast ranges from 80 °C to 100 °C (100 °C to 120 °C for CFL). Therefore, select the R_{on} of the MOSFETs so that no overheating occurs. In addition, always consider the MOSFETs packaging during selection. Generally, lower R_{on} values improve ballast performance at increased cost of the ballast. Always maintain losses in the MOSFETs to less than 1.5 % of the lamp power.

$$½\, \mathrm{I}^2{}_{coil} \times \mathrm{R}_{on} \tag{4}$$

Example:

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Ballast for a TLD 36 W lamp running at 40 kHz, where:

- LC = 1.9 mH, 8.2 nF
- I_{lamp} = 0.32 A (RMS)
- $I_{capacitor}$ = 0.20 A (RMS)
- I_{coil} = 0.38 A (RMS)

A pair of MOSFETs each with a R_{on} of 3 Ω when at operating temperature have a power loss of approximately 0.5 W (0.25 W per MOSFET).

4.2 Boot strap diode and capacitor

As both MOSFETs are N type, a voltage of approximately 12 V higher than bridge voltage (V_{bridge}) is required to drive the upper MOSFET. In addition, an external bootstrap capacitor is required to store the 12 V buffer charge. Figure 16 shows an example of an NXP Semiconductors half-bridge ballast controller IC with integrated bootstrap diodes that provide such a voltage. External capacitor C5 is used to store the buffer charge. Typically, a 100 nF, 50 V capacitor is sufficient for all similar applications.

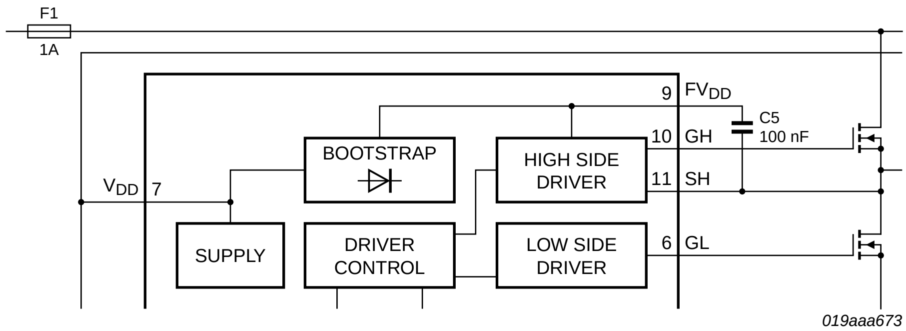

Fig 16. Boot strap

4.3 MOSFET gate drive RLC circuit

The gate drive circuit consists of a Resistor Inductor Capacitor (RLC) series circuit as shown in Figure 17. Resistance (R) is derived from the built-in resistance of the controller IC which is normally sufficiently high enough in most designs. However, an additional external resistor can be added if necessary. See Section 4.3.1.

Inductance (L) consists of parasitic PCB inductance coupled with the internal inductance inside the MOSFET. The capacitance (C) is the gate capacitance added with parasitic PCB.

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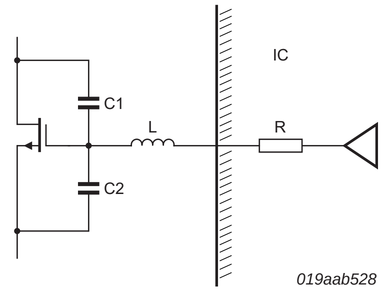
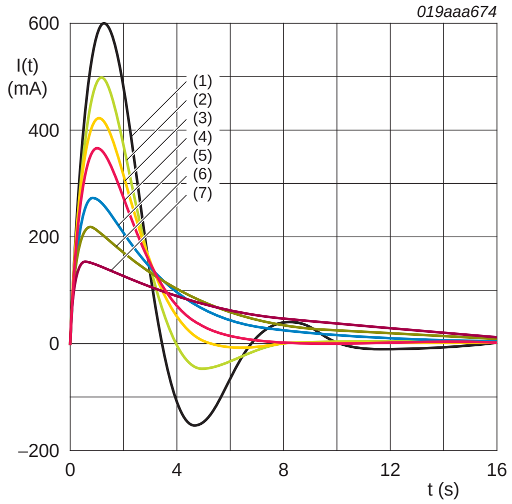


Fig 17. MOSFET gate drive circuit

For an RLC series circuit, the mathematical condition where no oscillation/undershoot in step responses exists is as Equation 5. See also Figure 18 line 4.

$$R^2 \times C^2 - 4LC >= 0 \quad (5)$$



(1) The critical damping plot is the bold red curve. The plots are normalized for L = 1 and C = 1

Fig 18. Under/overdamped responses of an RLC circuit

4.3.1 Extra resistance

As the L and C are known (given the layout), a minimum value of the series resistance (R) can be calculated. Where extra resistance is needed the extra series resistor (R) must be low enough for the gate drive not to open itself due to Miller capacitance. Hard switching is unavoidable at start-up. As an example, the upper MOSFET is switched on it must not lead to opening of the lower MOSFET due to the Miller capacitance. In such conditions, cross conduction occurs.

NXP Semiconductors controllers have internal resistors such that in almost all cases no additional external resistors are needed.

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Miller capacitance is known from the MOSFET specification, the inductor is the result of layout parasitic inductance of the MOSFET. Spice simulation and circuit measurements are the most effective method to achieve optimal component values. If it is not possible to find a resistor value that suits both conditions, then a resistor diode combination can be used. See Figure 19.

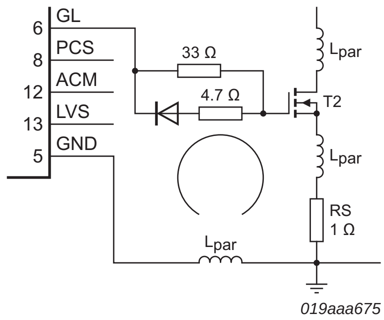

Fig 19. Extra resistance, lower MOSFET of half-bridge for example

The most important PCB layout parameters for a half-bridge are the loops of the MOSFET gate drives and the loop through the electrolytic bridge capacitor. The electrolytic bridge capacitor must be close to the half-bridge MOSFETs rather than close to the PFC MOSFET diode. Where this layout arrangement is not possible, add a 100 nF capacitor over the bus PFC voltage close to the half-bridge MOSFETs.

For more layout advice, refer to the SMPS layout and EMC guidelines.

5. Resonant tank

5.1 Introduction

The controller starts at its maximum frequency then quickly drops to the preheat frequency. The controller then waits for the preheat time to pass and then drops to the ignition frequency. Following ignition, the frequency of the half-bridge controls the power to the lamp. For example, dimmable controllers like the UBA2014, UBA2015(A) and UBA2016A all use a feedback control loop. The UBA2021 operates at a fixed frequency. See Section 3.4.

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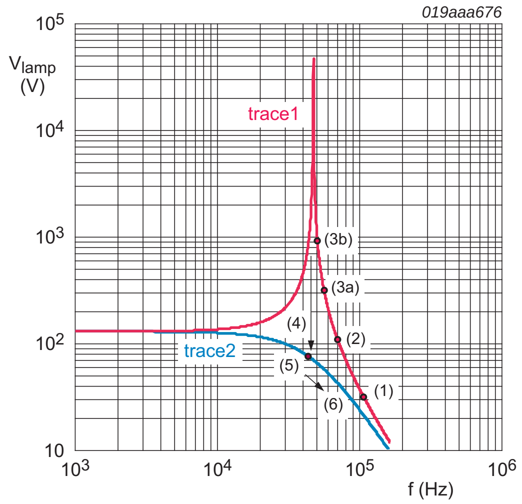
5.2 Frequency versus power



(1) Red line trace 1

(2) Blue line trace 2

Fig 20. Resonant tank frequency versus power

Key points for Figure 20

- 1 to 5 = Frequency sweep range
- 1 = Start
- 2 = Preheat (yes/no)
- 3a = Ignition warm
- 3b = Ignition cold
- 4 = Burn transition
- 5 = Burn
- 6 = Dimming

The resonant coil is the main dominating component of lamp power and the resonant capacitor an influential factor. The coil, resonant capacitor and lamp form the resonant tank. The coil and capacitor determine the resonant frequency.

In series resonant topology, as in Figure 1 the first selected component is the resonant capacitor. The resonant capacitor determines the current through the filaments during run time. In addition, it also determines the maximum preheat current given a maximum lamp voltage during preheat.

The more filament current that is needed the higher the resonant capacitor value must be. As an example, a value of 1.5 nF is practical for a small CFL and up to 10 nF for a TLD/T8 tube. A larger capacitor value also means that enough energy remains in the resonant tank for a single successful ignition attempt. For further details, see Section 5.3.

The next selected component is the resonant tank inductor (L) as shown in Figure 21. The lamp power depends on the frequency according to Equation 6.

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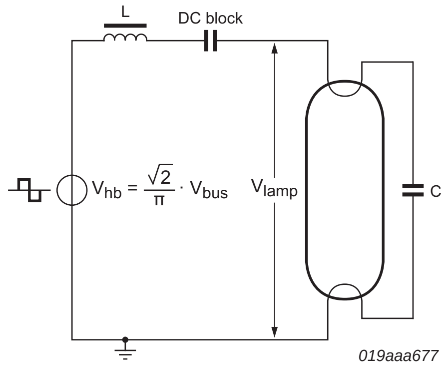

(1) See Equation 6

Fig 21. Resonant tank with a burner driven by a square wave voltage

$$P_{out} = \frac{V_{lamp^2}}{\omega L}\sqrt{\left(\frac{V_{bus}}{V_{lamp}}\frac{\sqrt{2}}{\pi}\right)^2 - (1-\omega^2 LC)^2} = \frac{V_{lamp}}{\omega L}\sqrt{\left(\frac{V_{hb}}{V_{lamp}}\right)^2 - (1-\omega^2 LC)^2} \quad (6)$$

Where: w = 2 × π × f

The same equation can be mathematically converted to determine frequency versus lamp power as shown in Equation 7.

$$f_{\%} = \frac{1}{2\pi}\sqrt{\frac{1}{LC} - 2\left(\frac{P_{\%}}{CV^2_{\%}}\right)^2 + \sqrt{\left[\frac{1}{LC} - 2\left(\frac{P_{\%}}{CV^2_{\%}}\right)^2\right]^2 - \frac{1-\left(\frac{4V_I}{V_{\%}\pi}\right)}{L^2C^2}}} \quad (7)$$

Where:

- L = Output stage inductor (H)
- C = Output stage capacitor (F)
- $P_{\%}$ = Lamp power at % dimming level (W)
- $V_{\%}$ = Lamp voltage amplitude at % dimming level (V)
- $f_{\%}$ = Frequency corresponding to lamp power at % dimming level (Hz)
- Vi = Bus voltage (V_{bus})

The DC blocking value is not accounted for in Equation 7, as a low blocking value has no significant influence over the results. More complex formulas that include the DC blocking capacitor/value (see Figure 21) into account can be found in Ref. 1

Figure 22 shows the power versus frequency graph using a series resonant PFC. In the example, L = 1.9 mH, C = 8.2 nF and V_{hb} = 400 V T8 36 W. Dimming at 10 % is reached before f_{max} and operation power is reached a few kHz higher than f_{min}.

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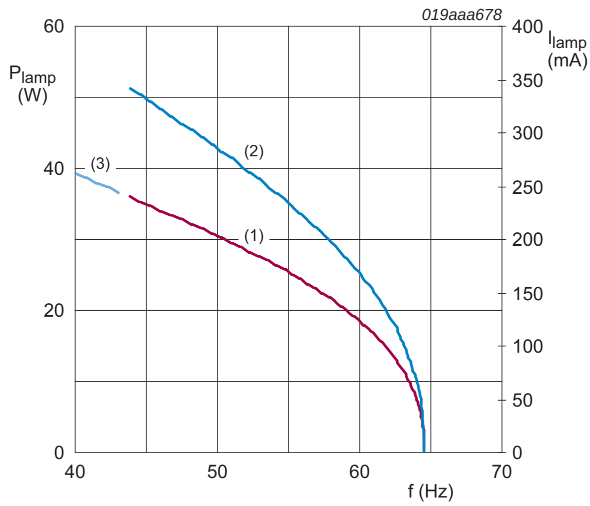

(1) P_{lamp}

(2) I_{lamp}

(3) $P_{fil} \times 1.5$

Fig 22. Power versus frequency

Table 1. Suggested resonant tank values for different lamp types

Lamp	Resonant coil L_{res} (mH)	Resonant capacitor C_{res} (nF)	f_{ph} (kHz)	R_{sense} (Ω)	V_{bus} (V)	I_{ph} (mA) (RMS)	V_{lamp} during preheat (V)	t_{ph} (s)
TLD 36 W	1.9 mH	8.2 nF	70 kHz	1 Ω	400 V	600 mA	230 V	1.7 s
TLD 58 W	1.4 mH	10 nF	60 kHz	0.8 Ω	400 V	700 mA	170 V	1.7 s
TL5 HE 14 W	3.9 mH	5.6 nF	54 kHz	3.3 Ω	400 V	225 mA	180 V	1.5 s
TL5 HE 21 W	3.7 mH	5.6 nF	54 kHz	3.3 Ω	400 V	225 mA	180 V	1.5 s
TL5 HE 21 W	4.0 mH	3.9 nF	58 kHz	3.3 Ω	400 V	225 mA	230 V	1.5 s
TL5 HE 35 W	4.0 mH	3.9 nF	58 kHz	3.3 Ω	400 V	225 mA	230 V	1.5 s
TL5 HO 39 W	2.0 mH	10 nF	52 kHz	1 Ω	400 V	560 mA	240 V	1.5 s
TL5 HO 49 W	2.6 mH	6.8 nF	52 kHz	2.2 Ω	400 V	370 mA	240 V	1.5 s
TL5 HO 54 ‘W	1.5 mH	10 nF	55 kHz	0.75 Ω	400 V	800 mA	330 V	1.5 s

5.3 Ignition and coil saturation

The resonant tank coil and capacitor value determine the preheat, ignition and operation conditions. A coils saturation current and losses are important parameters besides its value in Henries. The critical point for saturation is during lamp ignition. The reason is, the voltage over the lamp is at its peak, as is the current through the coil/resonant capacitor.

The required ignition voltage together with the chosen resonant capacitor and ignition frequency determine the level of current the coil must handle before saturation occurs.

Coil saturation causes high currents that damage the ballast. The UBA2015(A) and UBA2016A have direct coil saturation detection that increase the frequency and starts a fault timer in case of saturation. The UBA2021 increases its frequency and the UBA2014 jumps to its maximum frequency due to the capacitive mode that is a direct consequence

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of coil saturation. A ballast based on UBA2014 and UBA2021 can be extended with a small OverPower Protection (OPP) circuit. to bring them to standby in case of coil saturation.

The ignition frequency is the frequency where the resonant tank reaches the ignition voltage. Ignition frequency can be calculated using Equation 8.

$$F_{ign} = \frac{1}{2\pi\sqrt{\dfrac{(L_{res}(C_{res} \times C_{DCblock}))}{(C_{res} + C_{DCblock}) \times \left(1 + \left(\dfrac{V_{bridge \times 2}}{\pi \times V_{ign}}\right)\right)}}} \quad (8)$$

Where:

- V_{bridge} is the PFC output voltage
- L_{res} and C_{res} is the resonance coil/capacitor
- $C_{DCblock}$ is the DC blocking capacitor

The current through the coil as the resonant tank reaches the ignition voltage can be calculated using Equation 9:

$$I_{coil(max)} = V_{ign} \times 2 \times \pi \times F_{ign} \times C_{res} \quad (9)$$

The saturation value of L_{res} must be higher than the coil current during ignition. In addition, in cases of a restart/power dip the coil is still warm which can reduce its saturation current depending on the magnetic material used.

Under test, run a test with an aged burner as aged burners tend to have a higher ignition voltage than new ones. In addition, a cold burn has an increased ignition voltage. Finally, remember that the incoming photons accelerate tube ignition. Therefore the best stress test is an aged burner, shielded from any environment light straight from the freezer (leave 24 hours in between tests). The presence of a grounded metal plate close to the tube lowers the ignition voltage. Therefore, testing must take place under near to final design conditions.

No ignition protection

To protect the ballast TL controllers, the UBA2014, UBA2015(A) and UBA2016A all have lamp over voltage protection circuits. Lamp over voltage protection circuits limit the frequency ramp down during the ignition stage if the lamp voltage exceeds a level set by the external components. Figure 23 shows an example circuit using the UBA2014. The circuit prevents a high voltage over the lamp avoiding damage to the resonant capacitor, coil or other components. Following a time-out of about 100 ms the controller switches to standby.

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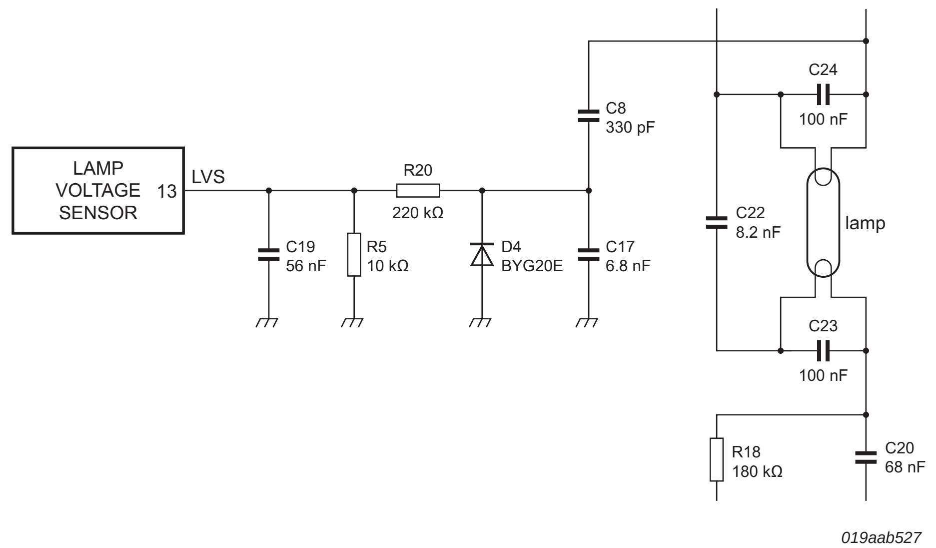

Fig 23. Lamp over voltage protection circuit

5.4 Preheat and run time filament current

The amount of preheat current versus time can be found in the burner specification as shown in Table 2.

Table 2. Preheat current versus time values

Lamp type PL-C 4-pin	RMS I_{ph} Preheat current (mA)	t_{ph}				R_{sub} (Ω)
		(0.5 s)	(1.0 s)	(1.5 s)	(2.0 s)	
10 W	min (RMS)	295	230	205	190	30 Ω
	max (RMS)	390	305	270	250	30 Ω
13 W	min (RMS)	300	240	215	200	30 Ω
	max (RMS)	395	315	280	265	30 Ω
18 W	min (RMS)	375	300	270	255	18 Ω
	max (RMS)	490	395	355	335	18 Ω
26 W	min (RMS)	560	450	405	380	9 Ω
	max (RMS)	735	590	535	500	9 Ω

The substitution resistor is 3.1 × cold filament resistance value during lamp operation. In addition, the burner specification gives maximum voltage during preheat without the risk of so called glow or early ignition. Glow is a discharge current without ignition which damages the filaments. When no burner specification exists, then as a guideline the filament resistance at the end of preheat must be 4.75 times the cold (room temperature) resistance.

The NXP Semiconductors controllers that have preheat, the UBA2021, UBA2014, UBA2015(A) and UBA2016A all have an input for a current sense resistor. Current sense resistors regulate the frequency so that the desired preheat current is obtained. The level of the lamp voltage under these conditions depends on the resonant tank LC components.

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During lamp operation, the current through the filament must be within limits as specified in the burner data sheet.

There are two kinds of specifications:

- The lead wire current versus discharge current.
- Sum of Squares (SoS) specification

5.4.1 Lead wire current versus discharge current

Figure 24 is an example of a specification of lead wire current versus discharge current. All currents are RMS values.

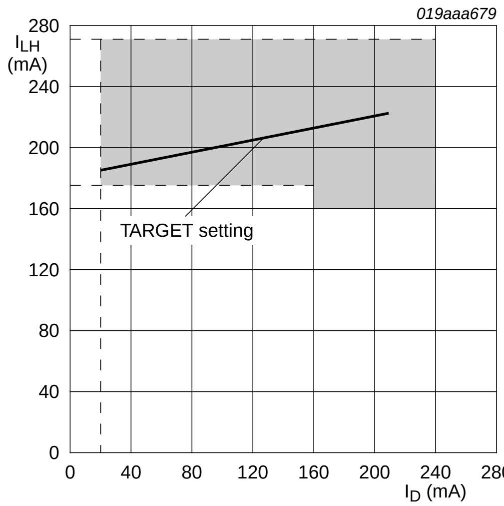
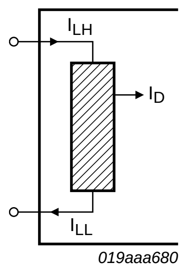


Fig 24. Example I_{LH} versus I_D burner specification



(1) I_{LH} = lead-high (total) current

(2) I_D = lamp (discharge) current

(3) I_{LL} = lead-low (heating) current

Fig 25. Definition of lamp currents

5.4.2 SoS specification

The specification can also be in the form of the so-called SoS of the I_{LH} and I_{LL} currents. SoS is the more recent and more accurate way of specifying filament currents. The equation for calculating the SoS value is shown in Equation 10.

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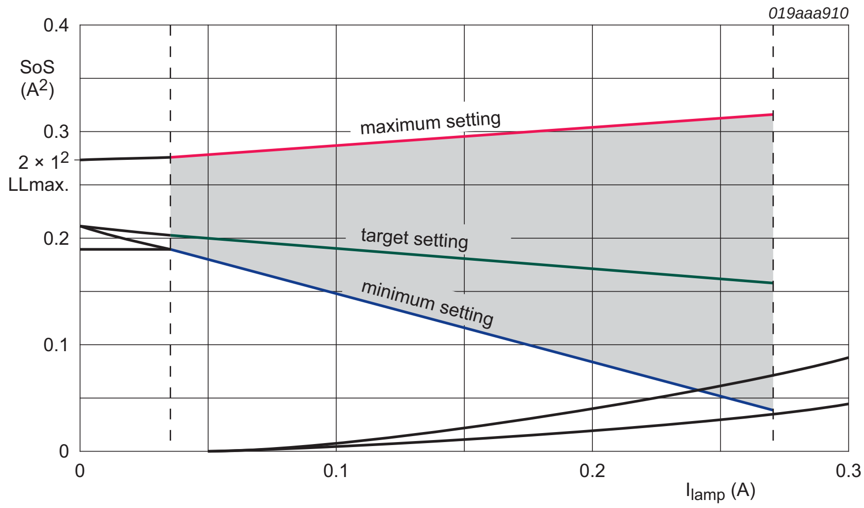

Fig 26. Graphical representation of the SoS lines of TL5 HO 24 W and 39 W

$$SoS = I_{LH}^2 + I_{LL}^2 \qquad (10)$$

6. Magnetic component design

6.1 Introduction

6.2 Inductor design parameters

An important factor of good ballast design is the quality of the main inductor. To achieve a high efficiency solution the inductance value, saturation current, proximity, core and ohmic losses, parasitic capacitance and stray magnetic fields are all important. Not understanding the functionality and implementing non-optimized components results in either, inferior performance or an impractical design. The following provides detailed guidelines for performance and design optimization.

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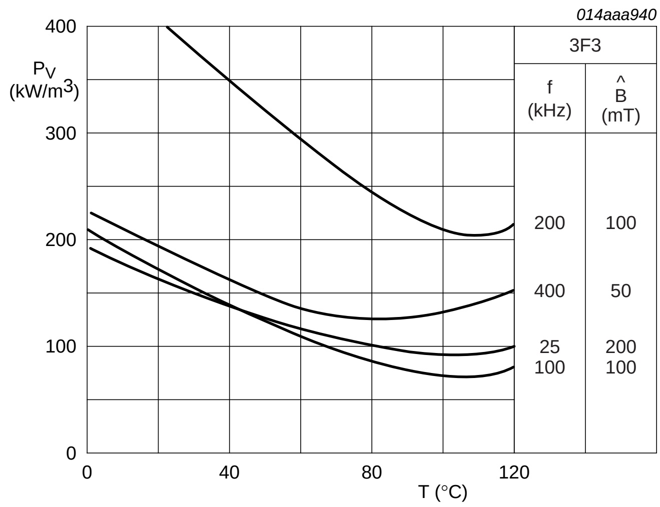

Fig 27. 3F3 specific power loss for several temperature/flux density combinations

For core material, each manufacturer has another code. For applications between 40 kHz and 100 kHz, 3F3 (Ferroxcube), N87 (EPCOS) or TP4 (TDG), are recommended. Select the material that has the lowest loss at working temperature. See Table 3.

Table 3. Ferrite core comparative geometry considerations

Aspect	Pot core/RM core	Double slab core	E core	Ec, ETD cores	PQ core	EP core	Toroid
Core costs	high	high	low	medium	high	medium	low
Bobbin costs	low	low	low	medium	high	high	none
Winding costs	low	low	low	low	low	low	high
Winding flexibility	good	good	excellent	excellent	good	good	fair
Assembly	simple	simple	simple	medium	simple	simple	none
Mounting flexibility	good	good	good	fair	fair	good	poor
Heat dissipation	poor	good	excellent	good	good	poor	good
Shielding	excellent	good	poor	poor	fair	excellent	good

6.3 Core type selection

Core geometry depends on several factors, for example, cost, flexibility, shielding and utilization factors. A core can have an inner core that results in a round or square winding shape. Stray inductance can vary with core shape. The maximum stored energy in the inductor together with the required air gap determines the core size.

A core with a large air gap can store more energy than a core with a small air gap. In practice, for resonant converters, an optimum design is reached when the core losses and the winding losses (proximity and skin losses) are balanced. Therefore it is necessary to compromise between high storable energy levels, low leakage inductance and small tolerances on the inductance. The maximum energy stored in the inductor can be calculated using Equation 11:

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$$E = \frac{1}{2} \times L \times I_p^{\ 2} \tag{11}$$

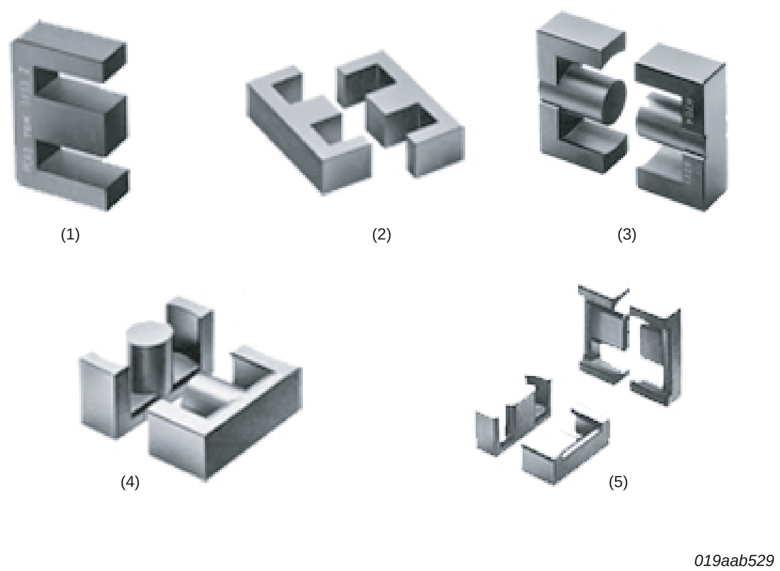
Example:

- L = 357 μH
- I_p = 1.48 A
- E = 3.9 $\times$ 10−4 J

The E series cores are the most commonly used for TL applications. CFL applications sometimes use different shapes for a mechanical fit. Table 4 and Figure 28 lists and shows some of the core types.

Table 4. Core selector

Core type	Typical size T5 HE 35 W
E	E20
EE	EE30
EER	EER25.5
EI	EI25
EPC	EPC27
ETD	ETD19



(1) EI core
(2) EE core
(3) EER core
(4) ETD core
(5) EPC core

Fig 28. Core types

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6.4 Calculate windings

A_l is often specified in the data sheet of the core material. It relates to the inductive value of a single turn on the selected core. Using this figure and knowing the inductance the number of inductor windings can be calculated using Equation 12:

$$N_L = \sqrt{\frac{L}{A_I}} \quad (12)$$

Example:

- A_l= 630 nH
- L= 357 μH
- N_L = 24

A practical value for N_L can be obtained by rounding the calculated value to its nearest integer. As a check, the magnetic material determines the maximum magnetic B-field. The peak value of B-field reached during operation has a substantial impact on core losses. Generally, the max B-field in the magnetic material must remain lower than the specified saturation field B_{sat} of the material. The B-field can be calculated using Equation 13:

$$B_{max} = U_e \times \frac{N_{L3} \times I_p}{I_e} \quad (13)$$

Example: refer to Equation 14:

- N_{L3} = 24
- I_p = 1.48
- u_e = 342
- l_e = 35.6

$$B_{max} = 342 \times 24 \times \frac{1.48}{35.6} = 338mT \quad (14)$$

6.5 Auxiliary winding count

CAUTION

Check that good isolation exists between the primary and secondary windings. As during ignition, a voltage difference of up to 1.5 KV can occur.

The auxiliary winding can be used for the following three purposes:

- To heat the filaments
- To detect lamp failure
- Generate the required voltage to power the controller

Equation 15 and Equation 16 apply:

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$$V_{sec} = \frac{L_{sec}}{L_{prim} \times V_{prim}} \quad (15)$$

$$\frac{L_{sec}}{L_{prim}} = \frac{N_{sec}^{2}}{N_{prim}^{2}} \quad (16)$$

6.6 Select wire diameters

Wire diameter selection is a trade-off between available winding area, ohmic, proximity and skin losses. The use of wire gauge sizes smaller than 1 mm in diameter at operating frequencies lower than 100 kHz results in negligible skin losses. Wire gauge sizes larger than 1 mm in diameter, it is recommended that litz wire or multiple strands are used.

Skin depth can be calculated using Equation 17.

$$\delta = \sqrt{\frac{2 \times \rho}{2 \times \pi \times f_{eff} \times u_r \times u_o}} \quad (17)$$

Where:

- $u_o = 4 \times \pi \times 10^{-7}$
- ρ = resistivity = 1.7×10^{-8} (copper)
- U_r (copper) = 1

Example:

- At 50 kHz sinusoidal current, using copper, the skin depth is 0.3 mm.

For the coil in a resonant tank ballast only take the first harmonic into account, as the current is close to a sine wave. For a coil (transformer) with secondary windings (inductive mode filament heating) a square voltage wave form is present on the windings. The secondary current (unlike the one from the main coil) therefore has high frequency components. For the wire type of these secondary windings, consider only the 3rd and 5th harmonic.

The peak current value is a dominant parameter when calculating the ohmic losses. Estimate the value by calculating the wire resistance and calculating the average power dissipation in the wire. Generally, the current density must be between 300 circular mills and 500 circular mills per ampere.

Table 5 shows wire sizes relative to current.

Table 5. Wire selection table

Diameter (mm)	Nearest AWG	Area (mm^2)	Area (circular mills)	DC res. (Ω/M)	Typical current level (A)
0.1	38	0.008	15	2.195	0.04
0.2	32	0.031	62	0.549	0.15
0.25	30	0.049	97	0.351	0.24
0.315	28	0.078	154	0.221	0.38
0.355	27	0.099	195	0.174	0.49

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Table 5. Wire selection table *...continued*

Diameter (mm)	Nearest AWG	Area (mm^2)	Area (circular mills)	DC res. (Ω/M)	Typical current level (A)
0.4	26	0.126	248	0.137	0.62
0.56	23	0.246	486	0.070	1.22
0.71	21	0.396	781	0.044	1.95
16×0.2	-	0.503	992	0.034	2.48
37×0.2	-	1.162	2294	0.015	5.73
61×0.2	-	1.916	3782	0.009	9.45

During lamp operation, the coil current is at its maximum around the 10 % to 15 % dimming level of the lamp current. This condition applies to a typical ballast operating at 60 kHz to 70 kHz.

6.7 Half-bridge transformer design for deep dimming

A low capacitive coupling from the primary to the secondary windings is used to reduce ElectroMagnetic Interference (EMI). In addition, stray current that interfere with the lamp current measurement are reduced.

As shown in Figure 32 the coil can also be a transformer with two small secondary windings. Typical values of these secondary windings are 0.5 μH for none dimmable (large capacitor in series) and 25 μH for dimmable applications (small capacitor in series). See Figure 32 for an example schematic.

It is important to maintain low parasitic capacitance for deep dimmable applications as follows:

- C_{stray} primary to secondary < 25 pF
- C_{stray} secondary to secondary < 15 pF

The breakdown voltage across the primary, primary to secondary and secondary to secondary windings must be higher than the maximum voltage the ignition protection circuit allows.

6.7.1 Suggested transformer construction

Figure 29 shows an example of suggested transformer construction:

- Winding 6-5 is secondary winding connected to the hot side of the lamp
- Winding 7-8 is secondary winding connected to the cold side of the lamp
- Winding 3-2 is the primary winding; pin 3 connects to switching half-bridge MOSFETs

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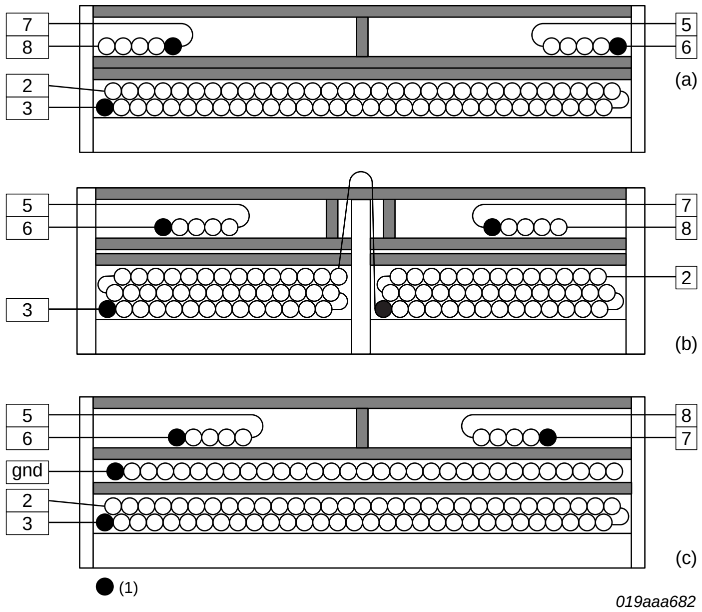

(1) Indicates mechanical start of the winding and electrical phase

Fig 29. Three different half-bridge transformer winding examples

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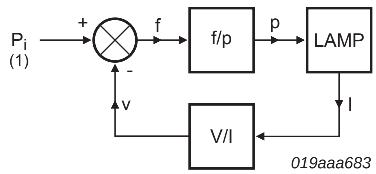
7. Dimmable controller feedback control loops

7.1 Introduction

Dimming ballasts need a feedback control loop to maintain the lamp current stable at the desired value. None dimmable ballast in principle can run at a fixed frequency. However, lamp current variations due to lamp temperature and aging can be avoided by using a controlled system. Figure 30 show an example of a feedback control loop.



(1) Desired lamp power

Fig 30. Feedback control loop

A desired lamp power is derived from the customer, including:

- 0 V to 10 V interface
- Dali
- Potentiometer
- Measurement of environment light
- Triac position for CFL ballasts

The UBA2014, UBA2015(A) and UBA2016A all have an input pin with a voltage range of 0 V to 2.5 V to set the desired lamp current. The topology and component values for a dimmable system can be different than for a non-dimmable.

7.1.1 Resonant tank values for a dimmable system

The frequency to lamp power/current takes place via the resonant tank as described in Section 5. It is important that the coil is designed with large losses at the deepest dimming level (highest frequency). A typical dimming curve is from 40 kHz at full power to 75 kHz at deepest dimming level.

7.1.2 Inductive mode heating

There are two commonly used filament heating topologies for fluorescent ballast controllers including:

- Classic series resonant, see Figure 31
- Inductive mode, see Figure 32

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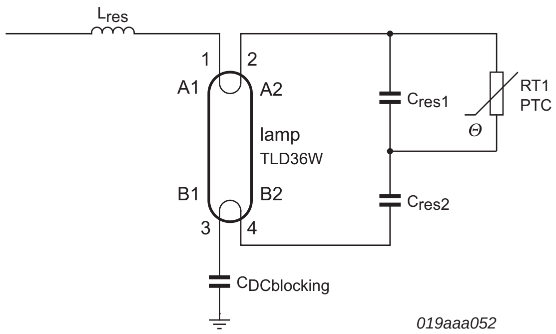

Fig 31. Series resonant heating

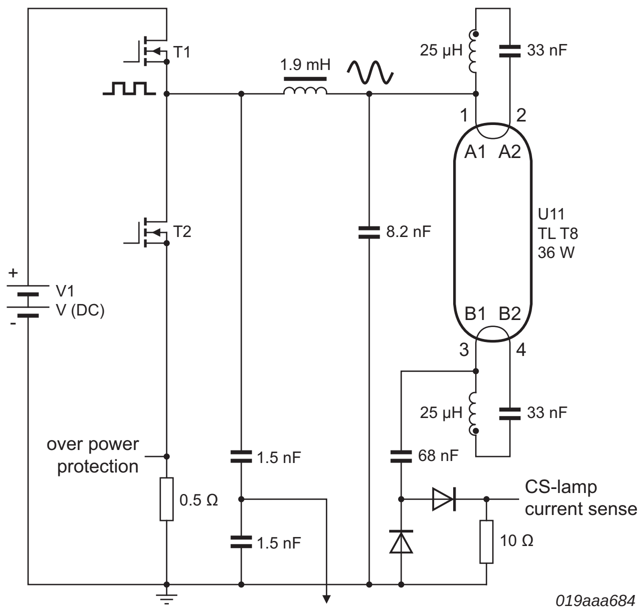

Fig 32. Inductive mode heating

The advantage of inductive mode heating is the lamp current can be measured separate from the current through the resonant capacitor. This advantage allows deeper dimming than with classic series resonant. The current through the sense resistor in the source of the lower MOSFET is the coil current which is the lamp current plus the capacitor current.

The deeper the dimming level the higher the resonant capacitor current becomes compared to the lamp current. For classic series resonant heating, at 15 % of nominal lamp current the feedback control loop regulates on the resonant capacitor current only. As a result it leaves the lamp switch-off. Therefore, dimming is limited to around 15 % see Figure 31.

The inductive mode topology allows dimming down to between 1 % to 2 %. To reach these deep dimming values, it is important to maintain low parasitic capacitance from primary to secondary on the coil and on the PCB.

Table 6 contains a list of values for different tubes for inductive mode heating topology.

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Table 6. Suggested resonant tank values for different lamp types

Lamp	Resonant coil	Resonant capacitor	Secondary inductance	Secondary capacitance	V_{bus}	Preheat energy	Preheat time
TLD 18 W	2.14 mH	4.7 nF	11 μH	39 nF	400 V	4 J	1.2 s
TLD 36 W	1.9 mH	8.2 nF	14 μH	33 nF	400 V	4.6 J	1.2 s
TLD 58 W	1.38 mH	8.2 nF	14 μH	27 nF	400 V	5 J	1.2 s
TLD 70 W	1.3 mH	8.2 nF	17 μH	27 nF	400 V	6.5 J	1.2 s
TL5 HO 24 W	2 mH	4.7 nF	10 μH	33 nF	400 V	3.2 J	1.2 s
TL5 HO 39 W	1.8 mH	4.7 nF	10 μH	27 nF	400 V	3 J	1.2 s
TL5 HO 54 W	1.3 mH	4.7 nF	8 μH	33 nF	400 V	3.8 J	1.2 s
TL5 HO 49 W	2.7 mH	4.7 nF	20 μH	15 nF	450 V	2.8 J	1.2 s
TL5 HO 80 W	1.1 mH	8.2 nF	15 μH	22 nF	400 V	5.8 J	1.2 s
TL5 HE 14 W	3.65 mH	4.7 nF	18 μH	22 nF	400 V	2.2 J	1.2 s
TL5 HE 21 W	3.56 mH	4.7 nF	18 μH	22 nF	400 V	2.2 J	1.2 s
TL5 HE 28 W	3.83 mH	4.7 nF	16 μH	22 nF	450 V	2.2 J	1.2 s
TL5 HE 35 W	3.88 mH	4.7 nF	16 μH	22 nF	480 V	2.2 J	1.2 s

7.1.3 Loop control speed

Figure 30 shows that the controller is a part of a feedback control loop (for all dimmable systems, and also for high end non-dimmable).

For deep dimming a fast feedback control loop is needed due to the physics of the lamp. If the lamp, does not receive power longer than 100 μs to 300 μs (ionization time constant) the discharge stops. Reignition is now needed.

NXP Semiconductors controllers only have on pin to set the feedback control loop speed. For systems with deep dimming two time constants are needed. One for the ignition sweep and average current setting and one to maintain discharge at low currents levels.

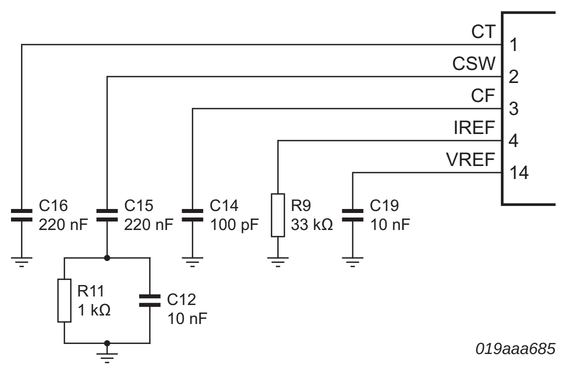

Fig 33. Fast feedback control loop

Figure 33 shows an example circuit using the UBA2014 with pin CSW configured as a fast control loop. Resistor R11 and C12 provides a small (a few kHz) and fast control around a base frequency determined by the voltage on capacitor C15. In Proportional Integral Differential (PID) controller terms, it is considered an extra pole.

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8. System performance

8.1 Introduction

Ballast losses are derived from:

- Losses in the PFC stage (typical efficiency for a designed PFC is around 96 % to 98 %).
- Losses in the switching MOSFETs (1 % to 2 %) see Section 4
- Losses in the resonant coil (1 % to 5 %) see Section 5
- Losses in the resonant capacitor (negligible if a high performance capacitor is chosen, however between 1 % to 2 % for low-cost capacitors)
- Miscellaneous losses in controller, diodes, sense resistor, and so on, 1 % in total

In total for a ballast optimized for performance the efficiency is around 93 %. A typical ballast optimized for Bill Of Materials (BOM) cost has an efficiency of approximately 85 %. The losses in the filaments are typically not considered to be ballast losses as heating the filaments is an essential part of the lamp operation.

8.2 Inductor losses

The inductor has several loss mechanisms. Calculation of these losses is complex and not clearly defined how these losses contribute to the overall total inductor losses. Equation 18 and Equation 19 simply illustrates a number of loss mechanisms within the inductor.

8.2.1 Ohmic losses

The combination of wire length and thickness causes ohmic losses. The calculation of the resistance and losses can be calculated with Equation 18 and Equation 19.

$$R_{DC} = \rho \times \frac{1}{A} \tag{18}$$

$$P_{DC} = \frac{1}{T} \times \int_{o}^{T} I^2 \times R_{DC} \qquad dt = \frac{1}{2} \times I_p^2 \times R_{DC} \tag{19}$$

Example, refer to Equation 20 and Equation 21.

- Wire length 1 m at diameter 0.56 mm
- ρ (copper) = 1.72×10^{-8}
- A = $\pi \times R2 = 2.46 \times 10^{-7}$
- I_p = 1.48 A

$$R_{DC} = (1.72 \times 10^{-8}) \times \frac{1}{2.46 \times 10^{-7}} = 70m\Omega \tag{20}$$

$$P_{DC} = \frac{1}{3} \times 1.48^2 \times 70 = 76.5mW \tag{21}$$

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8.2.2 Proximity losses

For proximity losses, the full calculations are outside the scope of this application note. However, it is important to understand that proximity losses are closely related to the skin depth and number of windings as shown in Figure 34.

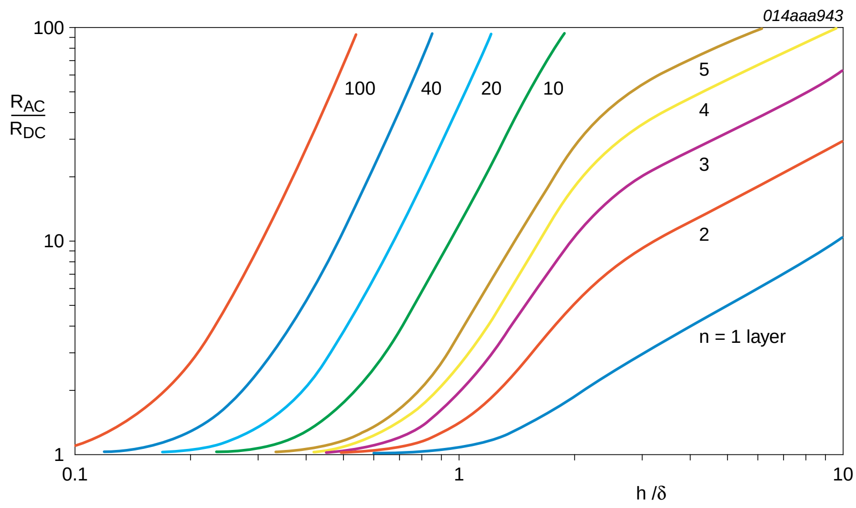

Fig 34. Proximity loss graph

Avoid too many layers of wires with radius that is close to, or below skin-depth. Normally, the proximity losses are calculated as a factor of the DC wire resistance as shown in the Equation 22.

$$R_{AC} = n \times R_{DC} \tag{22}$$

By maintaining low resistive losses, proximity losses remain minimal.

8.2.3 Core losses

The magnetization curve and frequencies determine the core losses in the magnetic material. At each converter cycle, the magnetic flux density excites the magnetic field in the core material. This result, produces a curve that is highly non-linear with the saturation level and hysteresis. The surface area enclosed by the variation in B-field strength at a certain frequency determines the losses. A bigger core, a higher B-field and a higher frequency increase these losses. The core material data sheet shows the loss per unit of volume at given frequencies. See Figure 35 and Figure 36.

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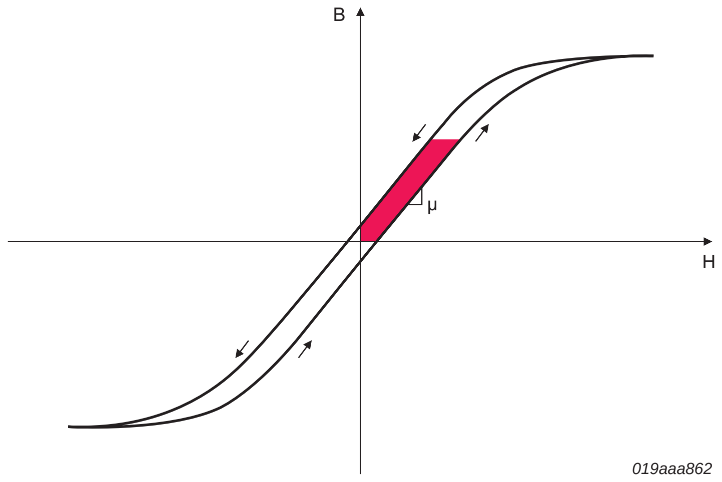

Fig 35. BH curve magnetic material

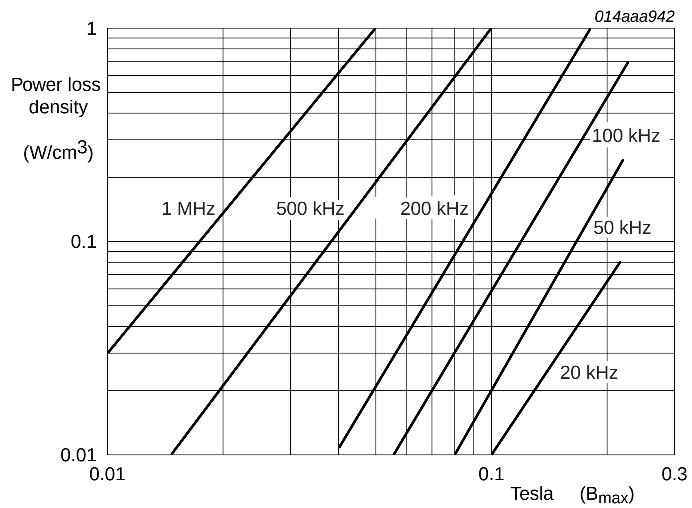

Fig 36. Material loss graph

A simple empirical formula that calculates core loss is called a Steinmetz equation as shown in Equation 23.

$$P_h = K_h \times f^{\alpha} \times B_{max}{}^{\beta} \times V_{core} \quad (23)$$

K_h and $\propto$ depend on the core material. Equation 23 can be improved by including the harmonics of a square waveform as shown in Equation 24.

$$P_{nse} = K_h \times (2f)^{\alpha} \times B_{max}{}^{\beta} \times (\delta^{1-\alpha} + (1-\delta)^{1-\alpha}) \times V_{core} \quad (24)$$

Where:

- d = duty cycle
- B_{max} = the peak flux density

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- f = the frequency of the fundamental
- V_{core} = the volume of the core

We can see that a higher frequency, a higher flux density, a smaller duty-factor and a bigger volume all increase core losses. A bigger core does not always reduce core losses; if the B-field is already low. The increase in volume counter acts the lower losses due to reduce flux density.

Example:

- d = 50 %
- K_h = 0.05
- f = 80 (kHz)
- $\propto$ = 1.84
- B_{max} = 0.1 T
- β = 3
- V_{core} = 2.4 cm^3

8.3 Sense resistor losses

For the sense resistor losses, the coil current can be calculated in the same way as the MOSFET losses. The lower MOSFET is switched on half the time. Sense resistor losses can be calculated using Equation 25.

$$½ \times I_{coil}^{2} \times R_{sense} \tag{25}$$

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9. Abbreviations

Table 7. Abbreviations

Acronym	Description
BOM	Bill of Materials
CFL	Compact Fluorescent Lamp
EMC	ElectroMagnetic Compatibility
MOSFET	Metal-Oxide Semiconductor Field-Effect Transistor
HF	High Frequency
LF	Low Frequency
LRC	Inductor Resistor Capacitor
PFC	Power Factor Correction
PID	Proportional Integral and Differential
RLC	Resistor Inductor Capacitor
RMS	Root Mean Square
RON	On Resistance
SoS	Sum of Squares
SMPS	Switched Mode Power Supply
TL	Tube Light
ZVS	Zero Voltage Switching

10. References

[1] **Lamp Model IEEE Transactions on Power Electronics Volume 20, No 5, Sept 2005. —** An Improved Design Procedure for LCC Resonant Filter of Dimmable Electronic Ballasts for Fluorescent Lamps.

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