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UBA2024B CFL ballast 100~120 V (AC) without voltage doubler

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Abstract	Application note for the NXP UBA2024B CFL driver running on 100~120 V (AC) mains without a voltage doubler circuit

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1. Introduction

This application note describes the design process of a CFL ballast for mains voltages from 100 to 120 V (AC) and therefore this document is to be considered as an addition to the existing application note AN10713 (18 W CFL lamp design using UBA2024 application development tool and application examples).

An application development tool has been made available to simplify the design of the lamp and the calculation of the resonance circuit and can generate a bill of materials needed to build the application.

The UBA2024 is a family of integrated half-bridge power IC's, designed for use in an integrated/sealed Compact Fluorescent Lamp (CFL) with lamp powers of up to 26 W. Typical input voltages are 100 V (AC) to 127 V (AC) and 220 V (AC) to 240 V (AC). The term lamp is used when the burner and electronic ballast are meant.

The UBA2024 includes both half-bridge power transistors with a level-shifter and drivers, bootstrap circuitry, an internal power supply, a precision oscillator and a start-up frequency sweep function for soft start and/or quasi-preheating. Due to the high level of integration, only a few external components are needed in a lamp ballast with the UBA2024.

The UBA2024 family of integrated CFL ballast controller IC's have different R_{on}, package and current rating, which is shown in Table 1 below.

Table 1. The UBA2024 family

Type	Package			Parameters	
	Name	Description	Version	$R_{DS(ON)}$	I_{SAT}
UBA2024P	DIP8	plastic dual in-line package 8 leads (300 mil)	SOT97-1	9 Ω	900 mA
UBA2024T	SO14	plastic small outline package 14 leads; body width 3.9 mm	SOT108-1	9 Ω	900 mA
UBA2024AP	DIP8	plastic dual in-line package 8 leads (300 mil)	SOT97-1	6 Ω	1350 mA
UBA2024AT	SO14	plastic small outline package 14 leads; body width 3.9 mm	SOT108-1	6.4 Ω	1200 mA
UBA2024BP	DIP8	plastic dual in-line package 8 leads (300 mil)	SOT97-1	2 Ω	2500 mA
UBA2024BT	SO14	plastic small outline package 14 leads; body width 3.9 mm	SOT108-1	2 Ω	2500 mA

Table 2 shows an overview of the possible applications for each member of the UBA2024 family, w.r.t. mains voltage and input configuration.

Table 2. The UBA2024 application range

Type	Lamp power[1]	Mains voltage	Input configuration
UBA2024P UBA2024T	5 W to 14 W	100 V (AC) to 127 V (AC) 220 V (AC) to 240 V (AC)	Voltage doubler Standard
UBA2024AP UBA2024AT	15 W to 18 W	100 V (AC) to 127 V (AC) 220 V (AC) to 240 V (AC)	Voltage doubler Standard
UBA2024BP UBA224BT	5 W to 26 W	100 V (AC) to 127 V (AC)	Standard

[1] Overall lamp power including driver circuit

1.1 UBA2024 family features

- Integrated half-bridge power-IC for CFL applications (both power and controller)
- Accurate oscillator with adjustable frequency
- Soft-start by frequency sweep down from start frequency
- Quasi preheat option (programmable sweep down timing)

1.2 System benefits

- Allows for very compact integrated lamp ballast which fits a small shell
- Low cost CFL applications due to low component count
- Higher reliability due to low component count
- Longer lamp life due to quasi preheat
- Easy applicable
- Based on EZ-HV SOI (silicon on insulator) technology
- UBA2024/UBA2024A can withstand a maximum voltage of 550 V
- UBA2024B can withstand a maximum voltage of 250 V

1.3 UBA2024B benefits

The half-bridge power transistors of the UBA2024B have a lower R_{on} and allow a higher current through the power transistors. However the breakdown voltage is limited and therefore a UBA2024B cannot be used for mains voltages higher than 127 V (AC).

To achieve operation with burner voltages of 80 V (RMS) and higher from a 100 to 120 V (AC) mains voltage two topologies are commonly used, as shown in Fig 1.

The first possibility is to use a so called "voltage doubler" circuit, for which an extra electrolytic capacitor is needed. On top of that the half bridge switches require a voltage rating equal to that for a 230 V (AC) application.

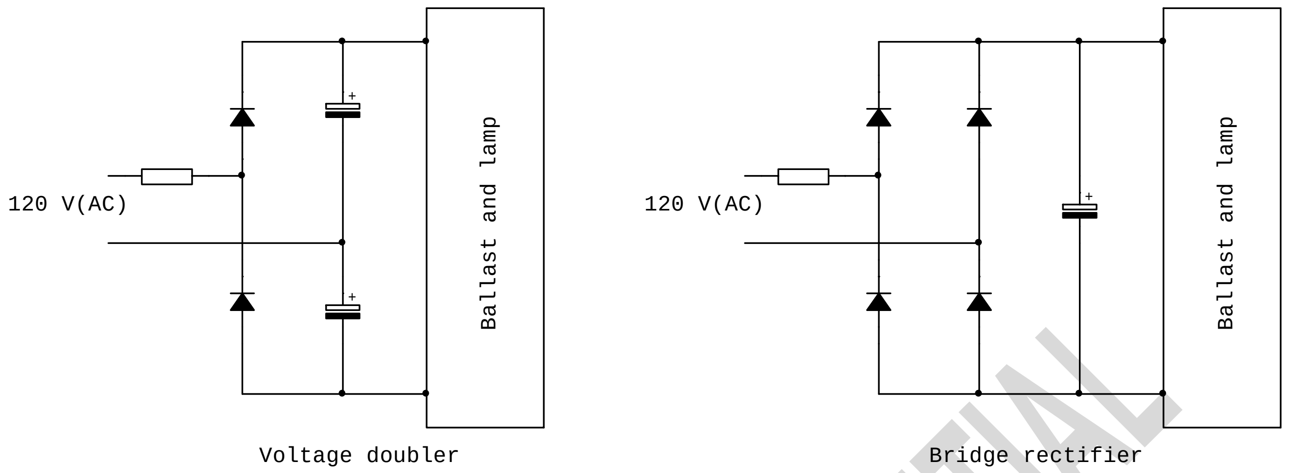


Fig 1. Mains input configurations for 100 V (AC) to 120 V (AC)

For more information about the voltage doubler topology please refer to application note AN10713 (18 W CFL lamp design using UBA2024 application development tool and application examples).

Another possibility to drive burners with voltages over 80 V (RMS) is a method using resonant gain from the LC-tank without a voltage doubler. This method is presented in this application note. The benefits are a lower voltage rating of the half bridge switches and no extra electrolytic capacitor for the voltage doubler.

When using resonant gain from the LC-tank the reactive current will be higher. As this current passes the integrated half bridge switches in the UBA2024B, these half bridge switches need to have a lower R_{on} to limit the power dissipation.

2. Circuit diagrams

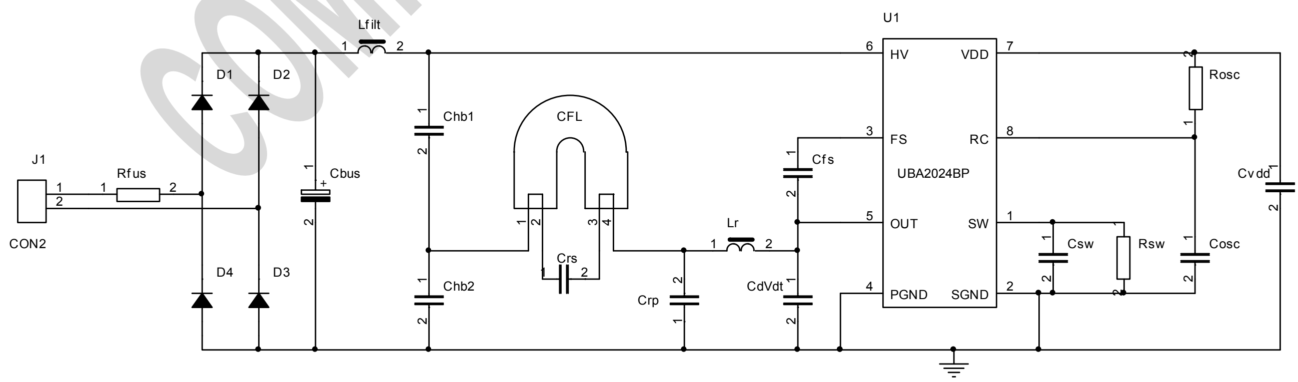


Fig 2. Application diagram for the UBA2024BP

Fig 2 shows the typical circuit diagram of the UBA2024BP in a DIL-8 package. Fig 3 shows a version with the UBA2024BT in an SO-14 package.

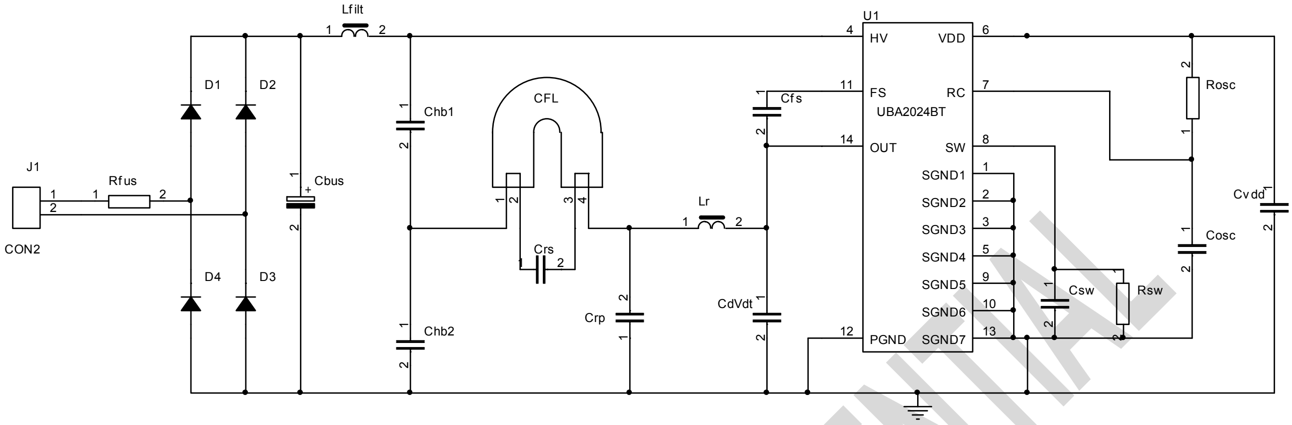


Fig 3. Application diagram for the UBA2024BT

The input circuit of the application comprises the fusistor R_{fus}, the diode rectifier bridge D1 to D4, and the buffer capacitor C_{bus}. L_{filt} suppresses harmonic disturbances to mains by the half bridge switching frequency. The controller IC is connected to timing components via pins RC for the oscillator and SW for the frequency sweep during preheat. The output of the IC drives the dV/dt capacitor, the resonant tank and the burner, where C_{hb1} and C_{hb2} are present for DC blocking. See the *UBA2024 datasheet* for a functional description of the IC.

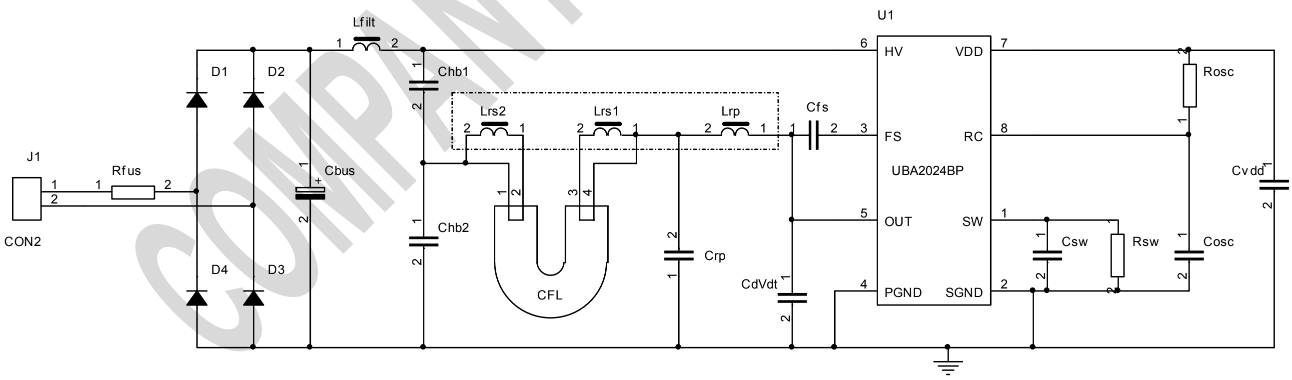


Fig 4. Application diagram for the UBA2024BP with inductive preheating

Additionally an example using inductive preheating with a UBA2024BP is shown in Fig 4. In this schematic only one resonant capacitor is needed, which is C_{rp}. In that case you can apply the total resonant capacitance here. The design of a circuit with inductive preheat lies beyond the scope of this document.

In Fig 2 and Fig 3 there are two resonant capacitors present, named C_{rp} and C_{rs}. In case the expected filament current (which in these two schematics is equal to the current through capacitor C_{rs}) is higher than the maximum allowed filament current I_{LL} the total resonant capacitance can be divided over both C_{rs} and C_{rp}. A part of the I_{LH} before C_{rp} was present will now pass through C_{rp} bringing I_{LL} to the desired value.

Fig 5 shows the flow of the lamp currents I_{LH} (Lead High), I_D (Discharge) and I_{LL} (Lead Low), where the discharge current is in fact the lamp current. The relation between these currents is as follows:

$$I_{LH} = \sqrt{I_D^{\ 2} + I_{LH}^{\ 2}} \qquad (1)$$

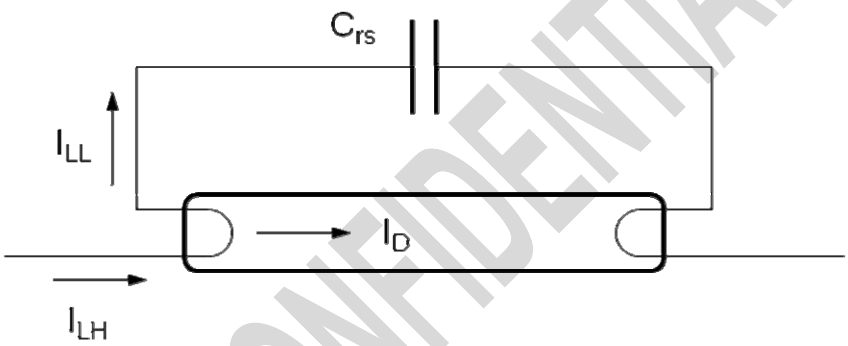


Fig 5. Lamp currents

So the total resonance capacitance amounts:

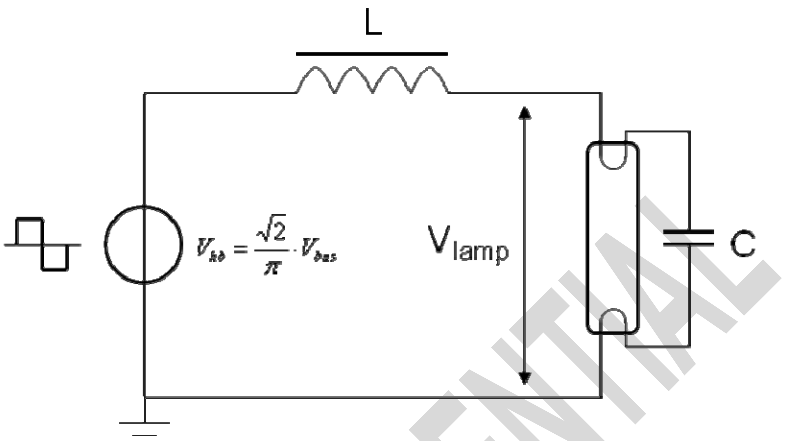
$$C_{res} = C_{rs} + C_{rp} \qquad (2)$$

Dividing the resonant capacitance C_{res} over C_{rs} and C_{rp} will result in the specified filament current and avoids decreased lifetime of the filaments in your burner by enhanced evaporation of the emissive material of the filaments and severe end-blackening of the tube.

3. Modes of lamp power control

First of all one should understand that a resonant tank with a self-inductance L, a capacitance C and a burner with an operating voltage V_{lamp} that is driven by a square wave voltage V_{hb} (the half-bridge output) with a given frequency f ($\omega = 2\pi f$) will result in a determined output power P_{out}. This is visualized in Fig 6.

Resonant gain (Q>1) is required when the burner operating voltage is higher than √2/π times the average bus voltage. Using resonant gain with a fixed frequency would give a very high dependency of the lamp power on the frequency and other component values. Therefore the spread in lamp power given normal component tolerances would be too high. An example of a transfer function with resonant gain running on a fixed frequency is shown in Fig 7. Also the resulting variation in power is shown when the frequency would deviate with ± 5 % around its nominal value.



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

Fig 6. Resonant tank with a burner driven by a square wave voltage

The UBA2024B can be used in two modes of operation, fixed frequency and frequency control by feedback. The choice of operation mode depends on the ratio between operation voltage of the burner and the bus voltage (rectified mains). Fixed frequency operation is applied when no resonant gain is required, which is the case when $V_{lamp} \leq V_{bus} \times \sqrt{2}/\pi$.

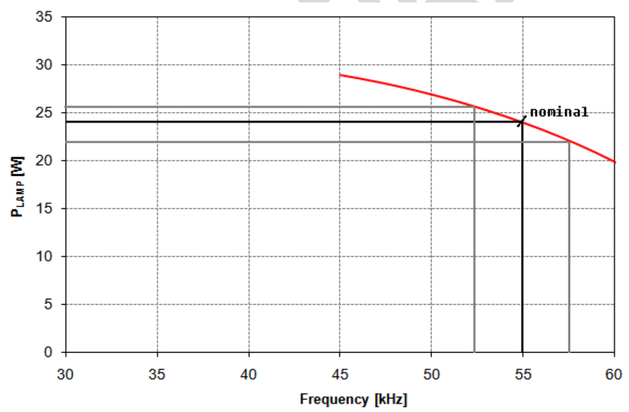


Fig 7. Resonant gain in a fixed frequency application with a resonant gain tank

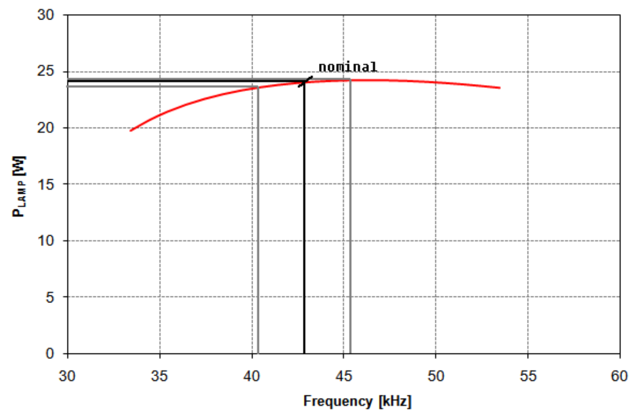


Fig 8. Resonant gain in a feedback controlled frequency application

A resonant tank driven close to its resonant frequency will behave almost similar to a current source. The lamp voltage will have a spread due to temperature, aging and production. Therefore in the case where a high gain is needed (as the lamp voltage is high) it is desirable to operate the tank close to the resonant frequency as this gives the smallest spread in power. This is shown in Fig 8, where a frequency deviation leads to only small variations in power. More details about this mode of operation can be found in Section 3.2.

First the fixed frequency operation is described.

3.1 Fixed frequency operation

Fixed frequency operation is well known and has proven itself with the UBA2024(A). The half bridge switching frequency is fully determined by R_{osc} and C_{osc} in the following expression:

$$f_{osc,HB} = \frac{1}{k \cdot R_{osc} \cdot C_{osc}} \quad (3)$$

The oscillator constant k has a typical value of 1.1, see the UBA2024 datasheet. The calculation tool calculates the inductor and capacitor values of the LC-tank in such a way that the IC will not run into hard switching at normal operation. For this mode of operation practical values for R_{osc} range from 50 kΩ to 400 kΩ. Note that the lower the value of R_{osc} is the higher the VDD output current is going to be, thus increasing the total package dissipation. Practical values for C_{osc} range from 100 pF to 1 nF. The advised value for C_{osc} is 180 pF for 40 kHz to 50 kHz and 270 pF for 25 kHz to 30 kHz.

The oscillator starting frequency is about 2.5 times the nominal frequency and gradually decreases, depending on lamp type and temperature, until the nominal frequency is reached. The lamp inductor L_r and the lamp capacitors (C_{rs} + C_{rp}) boost the lamp voltage gradually higher as the output frequency gets closer to their resonance frequency, until it is sufficient to ignite the lamp. In the meantime the current in the resonance circuit flows through the filaments providing quasi-preheating. The UBA2024 circuitry stops the frequency sweep at the resonance frequency, f_{res}, if the lamp has not ignited yet (see the UBA2024 data sheet for details). This ensures a maximum effort to ignite the lamp. The resonance frequency depends on L_r and the total capacitance C_{rs} and C_{rp}:

$$f_{res} = \frac{1}{2\pi\sqrt{L_r(C_{rs} + C_{rp})}} \quad (4)$$

As the ignition frequency, f_{ign}, is higher than or equal to the resonance frequency the resonance frequency should be chosen so that the preferred ignition frequency amounts $1.6 \times f_{burn} \le f_{ign} \le 1.8 \times f_{burn}$.

3.2 Feedback controlled frequency operation

We advise to use this topology when the burner operating voltage is higher than √2/π times the average bus voltage. The resonant tank needs to boost the voltage therefore the Q factor of the tank must be higher than 1 after the lamp has ignited (Q>1). The expression for the output power of the resonant tank is as follows:

$$P_{out} = \frac{V_{lamp}^{\ 2}}{\omega L} \cdot \sqrt{\left(\frac{V_{bus}}{V_{lamp}} \cdot \frac{\sqrt{2}}{\pi}\right)^2 - \left(1 - \omega^2 LC\right)^2} \quad (5)$$

Where

$$\omega = 2\pi \cdot f \quad (6)$$

The aim is to find inductor and capacitor values that result in the desired output power and at the same time direct the IC to the frequency where this power is reached. A transfer function with resonant gain will have a peak at a certain optimum frequency. An example of such a transfer function is shown in Fig 9.

Another goal is to control the running frequency of the IC such that there is zero voltage switching *and* the ballast is operating at the peak of the transfer function where the calculated lamp power is delivered.

The frequency control is made such that the IC works close to the peak frequency of the transfer function which has the benefit that at this point the slope of the transfer function is not very steep.

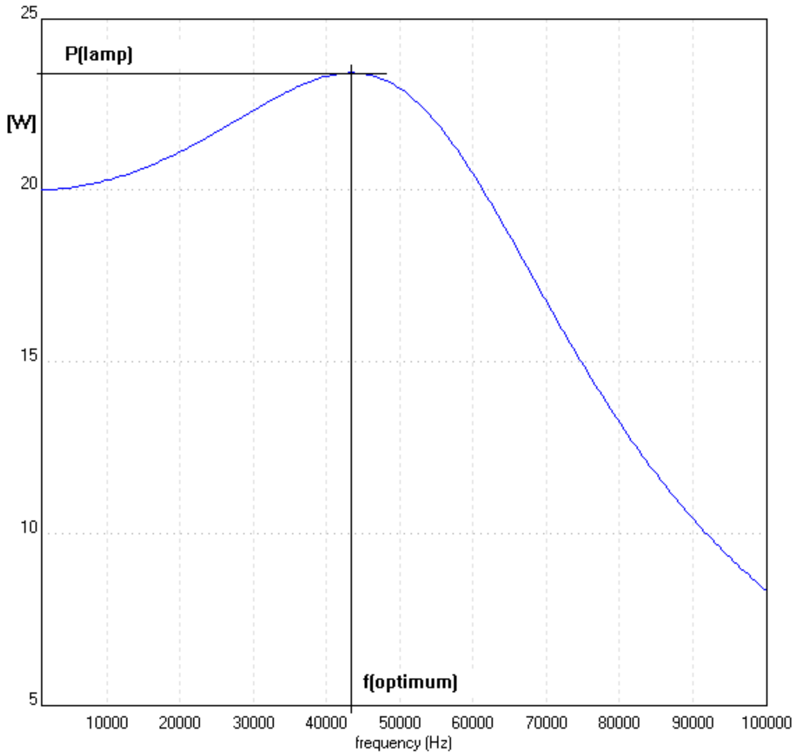


Fig 9. Example of a transfer function for a resonant gain LC-tank loaded with an ignited burner

3.2.1 Feedback controlled frequency using $C_{dV/dt}$

When an electronic ballast is running near capacitive mode (at Zero Voltage Switching) it means that the current through the LC-tank and lamp has a phase angle with respect to the half bridge voltage that is negative. In other words, the half bridge voltage lags the coil current. When the LC-tank and the lamp have an inductive character, the opposite is the case and this means that the coil current lags the half bridge voltage.

Close to the peak of the power transfer characteristic the phase shift of the coil current compared to the half bridge voltage will be very low, as shown in Fig 10. We use the coil current with a calculated $C_{dV/dt}$ such that the UBA2024B will operate on the edge of hard switching.

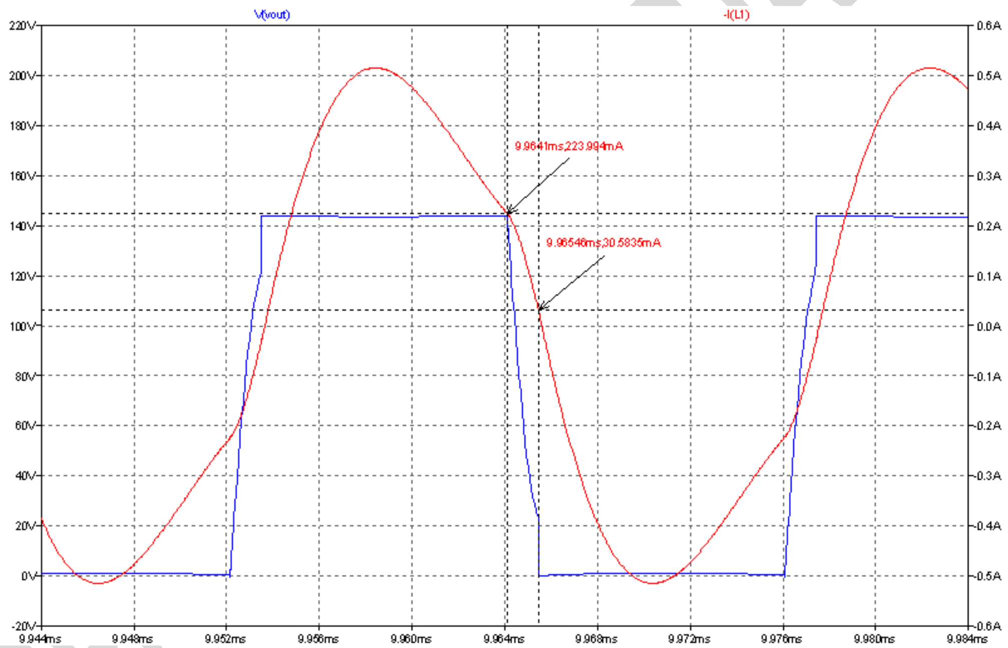


Fig 10. Half bridge output voltage and coil current with $C_{dV/dt}$ controlled frequency

The UBA2024 has a feature inside, which is originally meant for self protection during ignition. In case the IC runs into a hard switching condition, the IC's internal self protection circuitry will internally draw charge out of the C_{sw} capacitor and as a result the switching frequency will increase. The frequency increase will reduce the hard switching to below 14 V and as a result the IC will not be damaged or overheated by the switching losses due to charging and discharging $C_{dV/dt}$. This internal self protection circuitry together with a calculated $C_{dV/dt}$ is in fact used as a feedback control loop to control the switching frequency.

At first the resonant tank and expected operating frequency are determined. In principle the operating frequency is a user input with an advised value of 40 kHz, but any frequency between 20 kHz and 80 kHz could be given. The NXP tool calculates the LC tank for a specified operating frequency that lies 2 kHz above the peak in the resonant tank transfer function. Given this condition and expression (5) an optimal resonance capacitance and inductance is found.

Another point is to make sure that the IC will run on this specified frequency and that it keeps running on this frequency. Be it that this is not a fixed frequency, it will vary a few kHz with component values and burner voltage (e.g. in case of a cold burner). This feedback control loop is achieved by calculating the capacitance for $C_{dV/dt}$ which is needed for a coil current during the dead time after the trailing edge of the half bridge output voltage that equals:

$$C_{dV/dt} = \cdot \frac{I_{deadtime} \cdot t_{deadtime}}{V_{bridge} - 1} \qquad (7)$$

The value of $C_{dV/dt}$ that will give the frequency of 2 kHz above the peak in the resonant tank transfer function depends on the components in the LC-tank and the properties of the burner.

In the case of frequency control by $C_{dV/dt}$ the coil current and phase determine the frequency. The dead time of the UBA2024B is fixed. With the coil current values at the start and the end of the dead time we can calculate a matching $C_{dV/dt}$ capacitor value such that the UBA2024B regulates the frequency to the frequency where the desired power is delivered.

When hard switching occurs, there is still a voltage present over the load with a certain polarity at the end of the dead time, because the coil current is still flowing into/from the load. This will lead to extra losses in the half bridge switches, but as stated earlier a protection feature will prevent excessive hard switching, thus the IC will operate near hard switching. The calculation tool calculates the needed $C_{dV/dt}$ that results in the allowed hard switch level at the wanted operating frequency.

This $C_{dV/dt}$ capacitor is charged and discharged by the inductive load during the dead time. The coil current must not change polarity before the other half bridge switch is switched on. Fig 10 shows the half bridge voltage and the coil current.

3.2.2 Feedback controlled frequency using zero crossing of the coil current

Fig 11 shows what happens when the tool returns a value for $C_{dV/dt}$ that is relatively small. This is the case when the LC-tank and burner combination already has a capacitive like character and does not need much additional capacitance to be running on the edge of hard switching at the operation frequency.

The coil current changes polarity during the dead time and at the end of the dead time the half bridge output is charged to the allowed level of hard switching by the opposite flowing coil current. The IC will also protect itself against this kind of hard switching by increasing the frequency. So this is another way of hard switching, caused by the zero crossing of the coil current during the non-overlap or dead time.

Frequency control at zero crossing of the coil current is also known as zero voltage switching, which is a technique known to be used in discrete Colpitts self-oscillating electronic ballasts.

Take notice that this is not a preferred method of running on the edge of hard switching. The slopes of the half bridge voltage are very steep which may cause EMI. A second disadvantage is that high currents are flowing through the body diodes of the half bridge switches which is disadvantageous for the efficiency. One could increase $C_{dV/dt}$ in order to achieve frequency control by $C_{dV/dt}$ (waveform as shown in Fig 10).

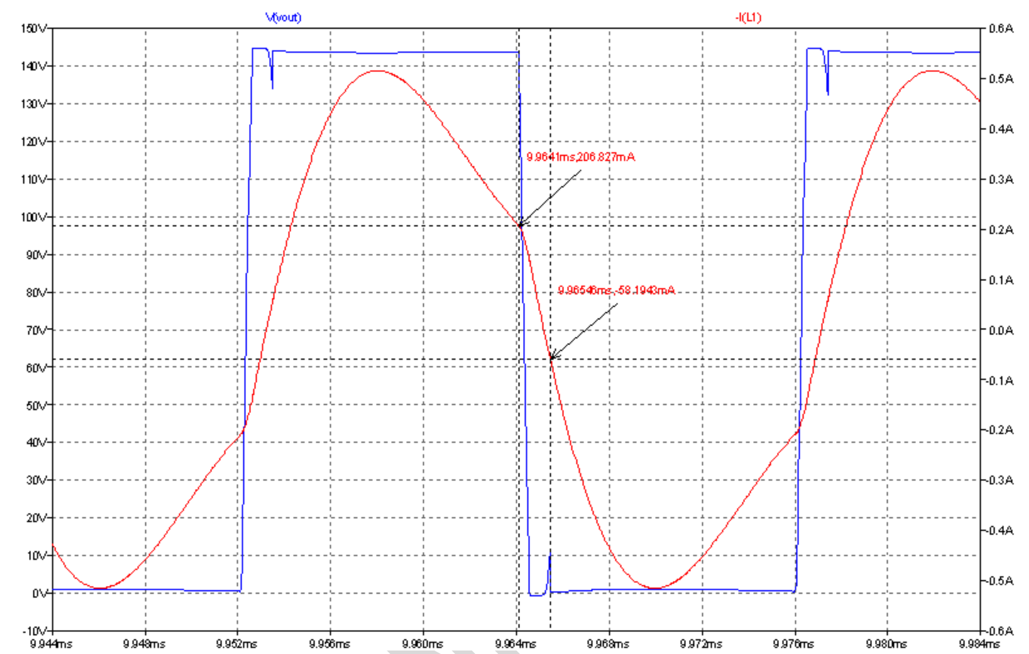


Fig 11. Half bridge output voltage and coil current with zero crossing controlled frequency

3.2.3 Losses due to hard switching

When the IC is working in the feedback controlled frequency operation, the IC will operate on the edge of hard switching, which will lead to additional losses. But the hard switching will not occur all the time, but as a function of the bus voltage ripple.

This is indicated in Fig 12.

When the IC is not hard switching the C_{sw} capacitor is charged and at hard switching the C_{sw} capacitor is discharged and the feedback system will balance itself. As a result the hard switching will only occur during approximately 25% of the time. The hard switching voltage will have a maximum level of 14 V and the average power losses due to hard switching can be calculated using expression (8):

$$P_{hsw} = \frac{t_{hsw}}{T_{V_{bus}}} \cdot f_{burn} \cdot C_{dV/dt} \cdot V_{hsw}^{\ 2} \qquad (8)$$

For the design example shown in Section 5 this would give less than 3 mW losses due to hard switching.

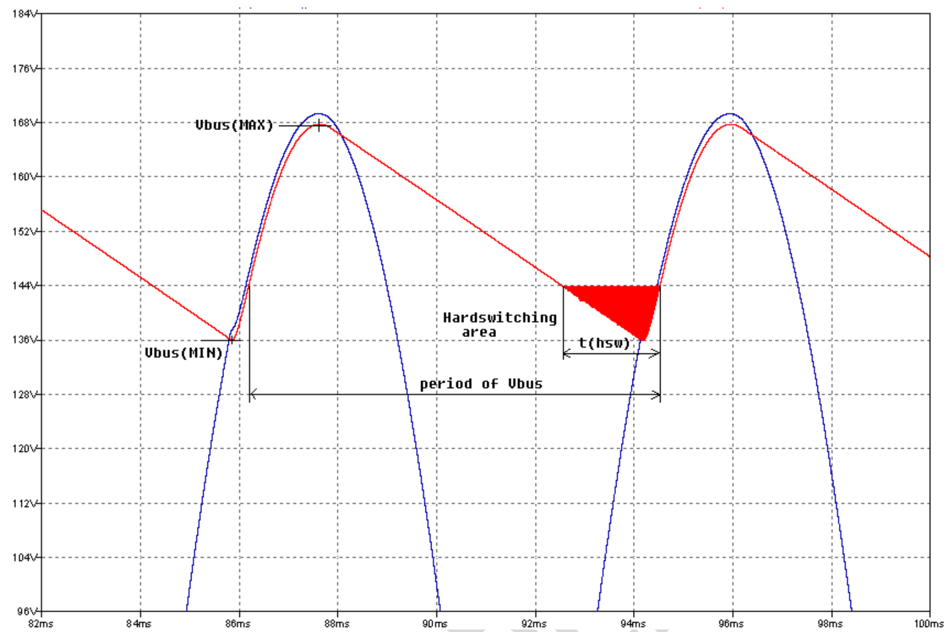


Fig 12. Hard switching occurrence

4. Preheating

In this chapter the preheat methodologies will be explained for a feedback controlled frequency application and for a fixed frequency application. For both topologies the starting frequency is set and as a consequence the time needed to reach the ignition frequency. Therefore the circuitry connected to the SW pin has changed compared to the default fixed frequency application as shown in the datasheet. The new schematic is shown below in Fig 13.

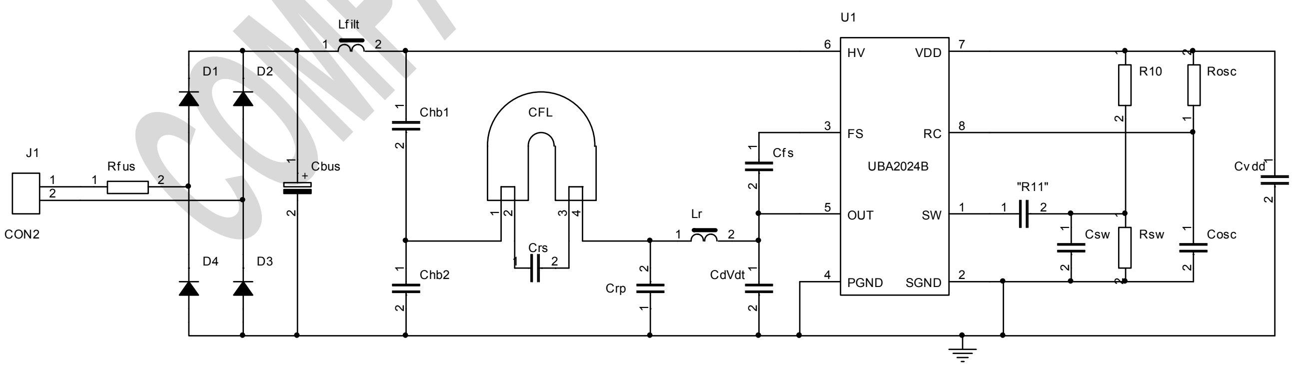


Fig 13. Schematic diagram of a fixed frequency application with new SW pin circuitry.

For realizing the applications as shown in Fig 2 and Fig 3 note that R10 is not mounted and that capacitor “R11” is replaced by a 0 Ω resistor.

It is not possible to have a controlled preheat current, in which case the current would look as shown in Fig 14. There is no free pin available and a sense resistor would lead to additional power losses.

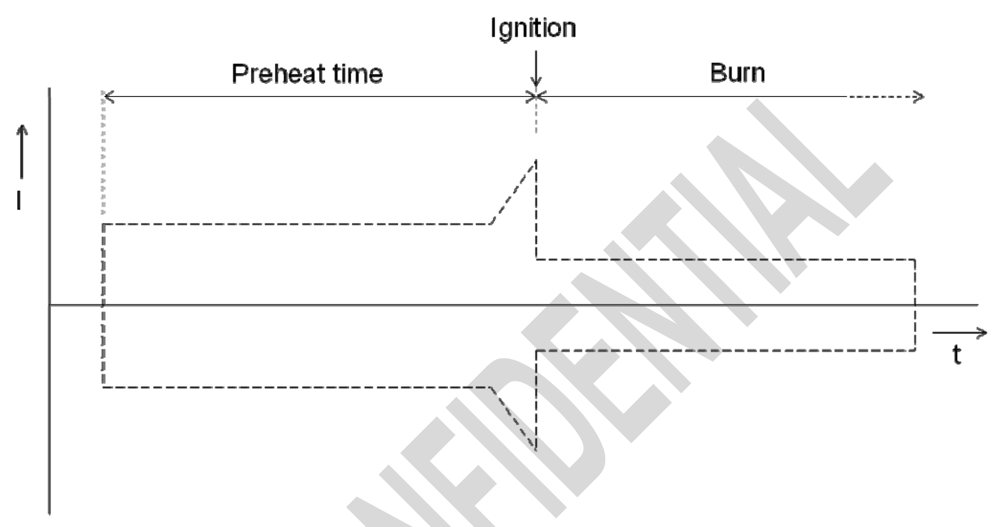


Fig 14. Controlled preheat current

The following sections of this chapter describe how a controlled preheat can be approximated for both a feedback controlled frequency application and a fixed frequency application. In Section 11.2 proof of concept is given that these approximations of a controlled preheat are adequate solutions to prevent lamp glow.

If filament specifications are not known, a rule of thumb is that the optimal ratio between the filament resistance at ignition and cold filament resistance is approximately 5. With a preheat time of 500 to 600 ms this ratio can be reached. With a cold start not only the ignition voltage is higher, also the starting voltage. Both ignition and starting cause more damage in case of cold start.

4.1 Start-up of a feedback controlled frequency application

Since the operating frequency is determined by operation on the edge of hard switching as explained in Section 3.2 the startup behavior of the application has been optimized to this mode of operation and the circuit that connects to the SW pin has changed compared to the default circuit as shown in the datasheet.

The timing components R_{osc} and C_{osc} are chosen in such a way that the oscillator starts at a desired preheat frequency, which typically lies approximately 10 kHz above the ignition frequency.

Now it is possible to set the starting frequency and as a consequence the time needed to reach the ignition frequency by choosing the right values for the timing components R_{osc}

and C_{osc}. The result of this method is that the designer in fact can program the preheat frequency and time. A major advantage of this method compared to a discrete solution is that one can give the same preheat energy as a discrete solution with a PTC resistor without this expensive PTC. Also a PTC resistor has to dissipate power to remain tripped during operation. Using a preheat time of at least 400 ms increases the switch cycle life time of the application and reduces the need for the saturation current through the coil as the ignition voltage decreases. The calculation tool will calculate R_{osc} for a given C_{osc} and a default preheat time of 600 ms and will return the actual preheat time. In case another preheat time is required the user can change R_{osc} and immediately see the effect on the calculated preheat time. In principle the minimum frequency is determined by the zero voltage feedback control loop, but to avoid problems with e.g. a cold burner a *safeguard* frequency is introduced by adding a resistor over C_{sw}, see Fig 15 below. The introduced R_{sw} will determine the minimum frequency of the oscillator.

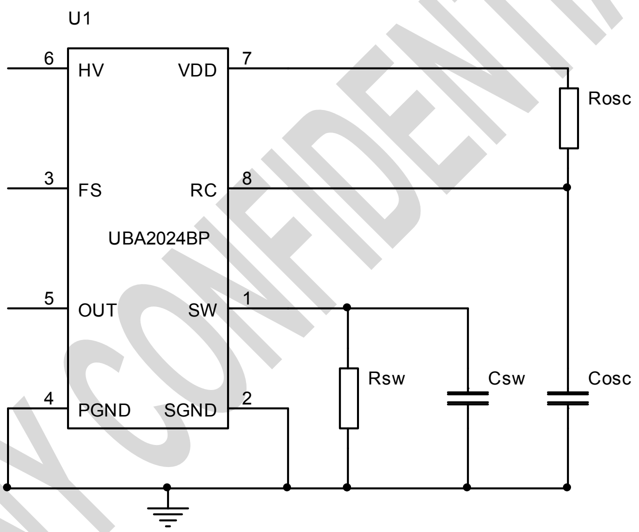


Fig 15. SW circuit for a frequency controlled feedback operated application

The voltage on the SW pin determines the amplitude and as a consequence the frequency on the RC pin. R_{sw} will limit the voltage on the SW pin, because normally C_{sw} will be charged with a current of 280 nA. At a level of 280 nA × 4.7 MΩ = 1.32 V C_{sw} will no longer be charged, hence the frequency will no longer increase. The time needed to reach this voltage is determined by C_{sw}.

Default values for R_{sw} and C_{sw} used in the calculation tool for a preheat time of 600 ms are 4.7 MΩ and 470 nF respectively.

In case of the fixed frequency operation applying the standard application from the datasheet, whereas the operating frequency is determined by the values of R_{osc} and C_{osc} the preheat frequency starts at 2.5 times the operating frequency, the preheat current as a function of time will look similar to the green dash curve in Fig 16, referred to as Quasi preheat starting at 100 kHz. However in the frequency controlled feedback operation where R_{osc} and C_{osc} only determine the starting frequency of the IC and $C_{dV/dt}$ determines the operating frequency, the preheat current will look similar to the solid blue curve in Fig

16. The advantage of the latter is that more energy is put into the filaments during the pseudo preheat which results in a more predictable ignition and an increased lifetime of the filaments.

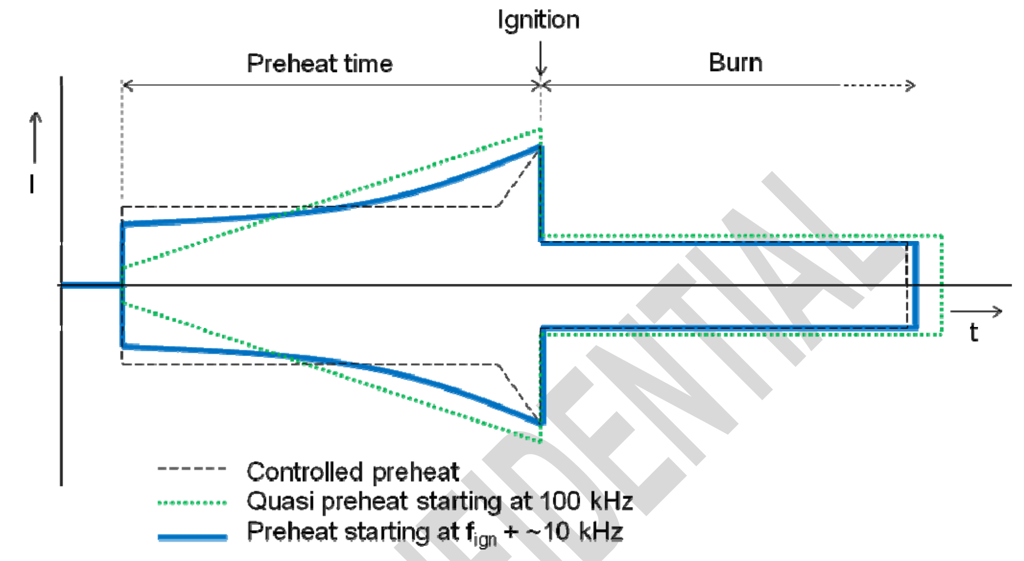


Fig 16. Preheat and ignition, preheat current as a function of time

In Fig 16 also a curve of the preheat current is shown of a system with controlled preheat (black dashed curve, already shown in Fig 14), where the frequency is constant during preheat and decreases to accomplish ignition after the preheat time has passed by. Note that preheating at a frequency that lies approximately 10 kHz above the ignition frequency results in a good approximation of controlled preheat system, e.g. Dragon.

For frequency controlled feedback operation the recommended value for C_{osc} is 1200 pF. Lower values of C_{osc} slightly decreases the duty cycle of the half bridge output and lead to higher hard switching losses on the leading edge of the half bridge output voltage. For fixed frequency operation, one can use smaller values for C_{osc} (see Section 4.2).

4.2 Start-up of a fixed frequency application

The time needed to sweep down (set by C_{sw} only, R_{sw} is not present when the IC is used in the standard application shown in the datasheet) from the start frequency to the resonance frequency can be used as an approximation for the ignition time. The sweep time is typically C_{sw} (nF) × 10.3 ms. For large values the ignition time is shorter, because the lamp ignites before the resonance frequency is reached. The typical ignition time is 1 s when C_{sw} = 330 nF. A larger C_{sw} increases the sweep time and improves the preheating of the electrodes. However, the rise of the pre-ignition lamp ignition voltage is also slower. Both a quasi-preheat that is too short and a voltage rise that is too slow increase the glow time of the lamp. This reduces the lifetime of the lamp. During the glow phase the lamp is ignited, but the filaments and the gas inside the lamp are not at their final operating temperature. The UBA2024 has a mechanism to push extra energy into the lamp during this glow phase, which is described in the UBA2024 datasheet. This will make the lamp go to its final light output quicker which gives a longer lifetime for the lamp. Typical values for C_{sw} are between 33 nF and 330 nF when the IC is used in the standard fixed frequency operation mode.

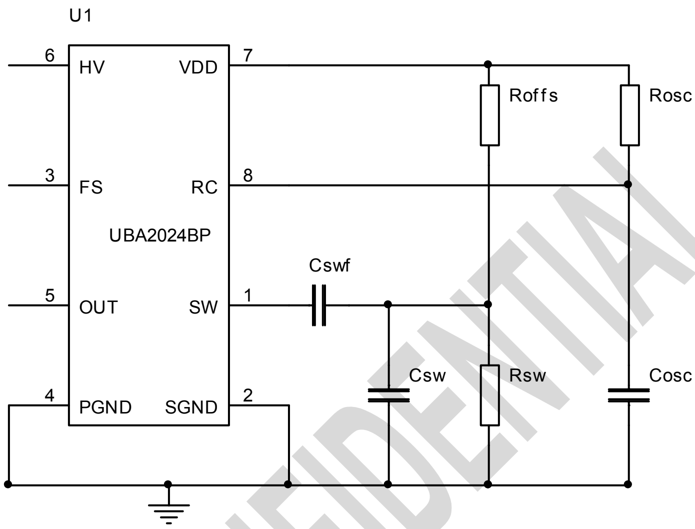


Fig 17. SW circuit for a fixed frequency operated application

In Fig 17 the schematic is shown of the SW circuitry which also provides a starting frequency that lies approximately 10 kHz above the ignition frequency. In this operation the operating frequency is still determined by R_{osc} and C_{osc} according to expression (4). The starting frequency is determined by the offset voltage that is determined by the voltage divider Roffs and Rsw. The capacitor Csw now works as a filter for this offset voltage. After start up Cswf will be charged further until the IC has reached the operating frequency. This preheating method is similar to the solid blue curve shown in Fig 16.

Default component values values used in the calculation tool are Cswf = 470 nF, Csw = 10 nF, Rsw = 10 kΩ and Cosc = 220 pF. These defaults are used by the calculation tool to determine E48 values for both Roffs and Rosc, resulting in a preheat time of 600 ms. Finally the tool will also return the actual preheat time using the calculated Roffs and Rosc.

5. Design of a 26 W non-dimmable CFL

This chapter explains the selection criteria for the component values. It also clarifies how to feed the application development tool with the appropriate values for components. With the calculation tool and with the help of some practical guidelines it should be easy to set up designs of different lamp powers. Throughout this document the light source itself is called the burner. The tool is meant for all use cases for burners operating at 50 V to 130 V of 8 W to 24 W. In this application note a PL-C 4P 26W burner which has a specified power of 24 W operating at 80 V is taken as an example.

5.1 Selecting a buffer capacitor and fusistor

Lamp power of a resonant tank with burner always depends on the bus voltage. For the use case of 220 V or 110 V (AC) with a voltage doubler this relation is more relaxed than for rectified 110 V (AC).

For proper operation a bus voltage ripple ratio, determined by the buffer capacitor, between 15% and 20% is recommended. If the buffer capacitor has a value resulting in a ripple ratio of less than 15%, the application will draw higher than necessary charge current peaks from the mains which will is disadvantageous for the power factor. In the tool this ratio is calculated and returned to the user. Choosing a smaller buffer capacitor will lead to a higher ripple and a lower average bus voltage. As a result of this the LC-tank may have to provide more resonant gain for which bigger resonant capacitors are needed. So choosing a smaller buffer capacitor does not necessarily lead to a smaller application. Also the Crest factor of the lamp power would become worse. The following table shows recommended values for the buffer capacitor and fusistor for a standard input configuration as shown in Fig 1 per power range of the application running on a mains voltage of 120 V (AC) and 60 Hz.

Table 3. Advised values for the standard input configuration

Lamp power range[2]	C_{BUS}	R_{FUS}[3]
4 W	10 µF / 200 V	18 Ω (0.5 W)
5 W to 6 W	15 µF / 200 V	12 Ω (0.5 W)
7 W to 8 W	15 µF / 200 V	12 Ω (1 W)
9 W to 11 W	22 µF / 200 V	5.6 Ω (1 W)
12 W to 14 W	22 µF / 200 V	5.6 Ω (2 W)
15 W to 18 W	22 µF / 200 V	5.6 Ω (2 W)
19 W to 22 W	33 µF / 200 V	3.3 Ω (2 W)
23 W to 26 W	33 µF / 200 V	3.3 Ω (2 W)

[2] Overall lamp power including driver circuit

[3] Minimum continous power rating

5.2 Using the calculation tool

This section describes how to use the calculation tool and how to interpret the results.

5.2.1 Input values

The application development tool calculates the component values based on the following input parameters:

- Burner power
- Burner operating voltage
- Burner ignition voltage
- Filament resistance

- Maximum filament current
- Mains input voltage and frequency (typical operating voltage)
- Combined value of the DC blocking capacitors

Fig 18 shows the part of the application development tool where the input parameters can be entered. The example shows the design of a 26 W lamp. This is the total lamp power, which means 24 W burner power and about 2 W loss in the electronic ballast. The burner used in this example is a replaceable burner. It is based on a G24q-3 fitting with the following parameters.

- Burner power = 24 W
- Burner voltage = 80 V
- Ignition voltage = 460 V
- Warm filament resistance = 9 Ω
- Maximum filament current = 320 mA

The following actions need to be taken:

1. Enter the burner parameters
2. Enter the mains voltage to be used for the 26 W lamp (120 V)
3. Enter the value of the buffer capacitor (33 µF)
4. Enter the mains frequency (60 Hz)
5. Enter the total value of the blocking capacitors (300 nF)
6. Enter the wanted operating frequency (40 kHz)

When using burners with an operating voltage up to 73 V the resonant tank does not have to provide resonant gain. When the UBA2024B is used with burners that have a high operating voltage the resonant tank has to provide resonant gain and the frequency should be regulated on the edge of hard switching. This frequency regulation is a protection of the UBA2024, which is meant for self protection during ignition. The advantage of resonant gain is that no voltage doubler capacitors are needed. These consume a lot of space in a retrofit CFL. The frequency at which the UBA2024B will run no longer depends on its RC timing components provided f_{min} is picked about 5 kHz below the actual run frequency. This will increase the accuracy of the system. The disadvantage of switching on the edge of hard switching is that there are small switching losses in the half bridge. For this application the additional switching losses amount less than 15 mW

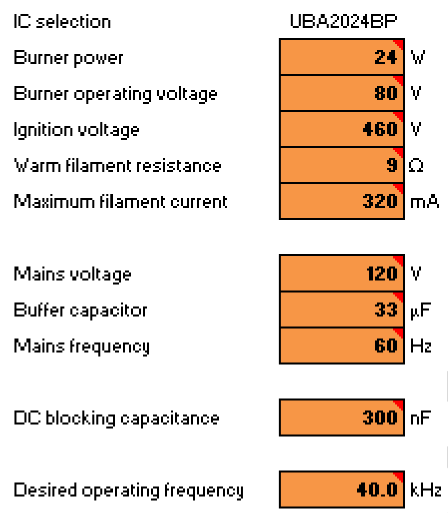


Fig 18. Entering the design parameters for a 26 W lamp

Based on the burner parameters, mains voltage and frequency, the buffer capacitor and selected DC blocking capacitors and the wanted operating frequency the calculation of the LC resonance tank can be executed by pressing the Optimize! button (Fig 23). The application development tool then calculates advisory values for the resonance inductor and capacitor and for the dV/dt capacitor. Also the actual operating frequency is calculated.

5.3 Calculation algorithm

After entering all the necessary parameters the calculation will proceed by returning advised values for the resonance capacitor and inductor. In case the calculated filament current is higher than the entered maximum value the resonance capacitor will be split up such that the requirement for the filament current is met. Instantaneously the capacitances will be returned as E12 values.

The next step in the calculation is that a zero voltage switching condition is achieved by determining the average coil current during the non-overlap time. The goal is to have no difference between the actual average coil current during the non-overlap time and a certain desired coil current during the non-overlap time. Now a value is determined for the dV/dt capacitor that leads to this condition. The found capacitance will also be returned as an E12 value.

The next step in the calculation is to fine tune to the zero voltage switching condition after the dV/dt capacitor value has been adapted to an E12 value. So, once again, the goal is to have no difference between the actual average coil current during the non-overlap time and a certain desired coil current during the non-overlap time. This is achieved by changing the actual operating frequency f_{burn} and the resonance inductor L_{res}, under the following constraints:

- Average lamp power equals the wanted burner power

- Actual operating frequency is less than or equal to the tank's resonant peak frequency increased with 2 kHz when the lamp is ignited.

The final step is to enter the RC timing components R_{OSC} and C_{OSC}. For C_{OSC} a value higher than 1 nF is recommended (default is 1.2 nF) and R_{OSC} is advised such that the preheat time equals 600 ms. Enter a realistic value that approximates the advised value and the tool will calculate the preheat time instantaneously.

5.4 Calculation results

Once the calculation is complete the tool will return graphs of the average burner power as a function of the rectified bridge voltage (Fig 19), burner power as a function of frequency (Fig 20) and burner voltage, filament current and frequency as a function of time (Fig 21) during start-up (preheat and ignition) of the application.

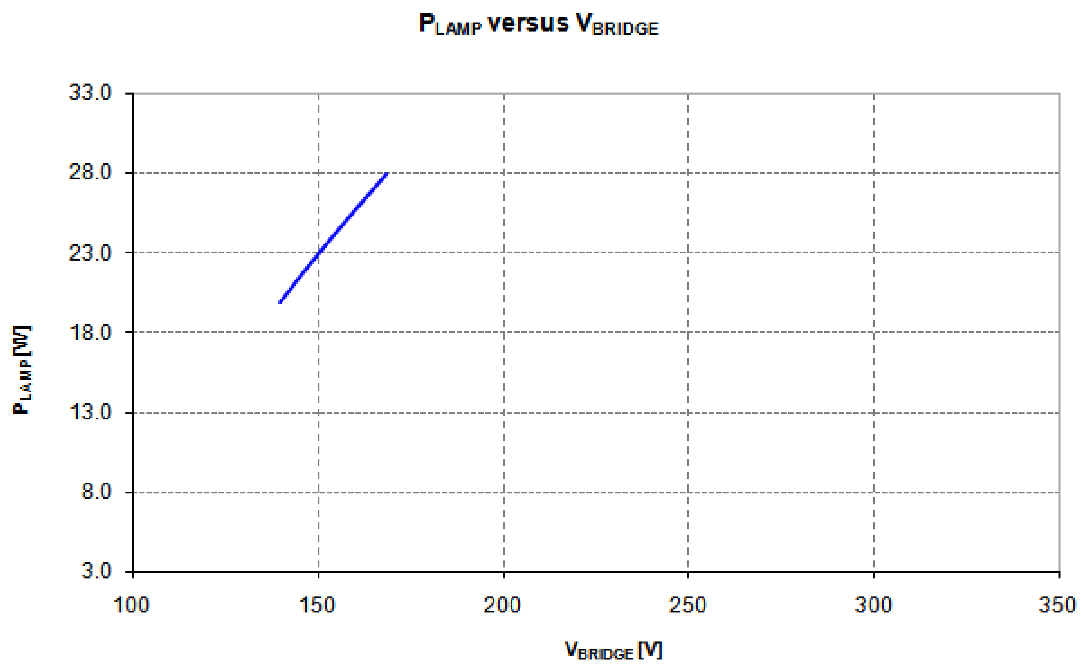


Fig 19. The average burner power as a function of the rectified bridge voltage

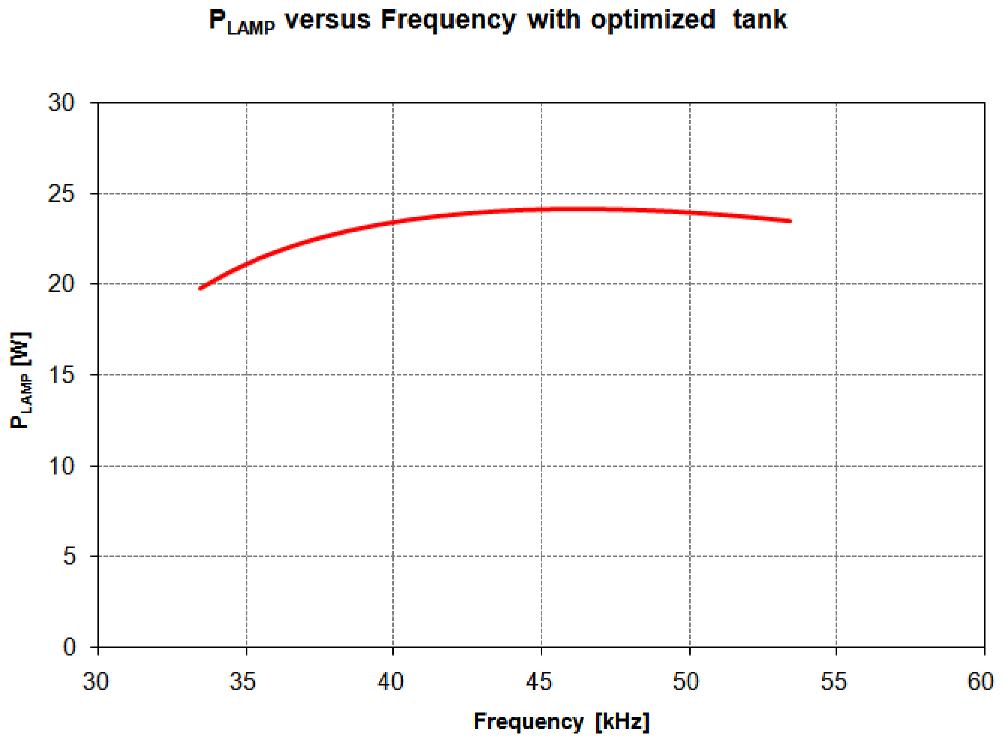


Fig 20. Burner power as a function of frequency

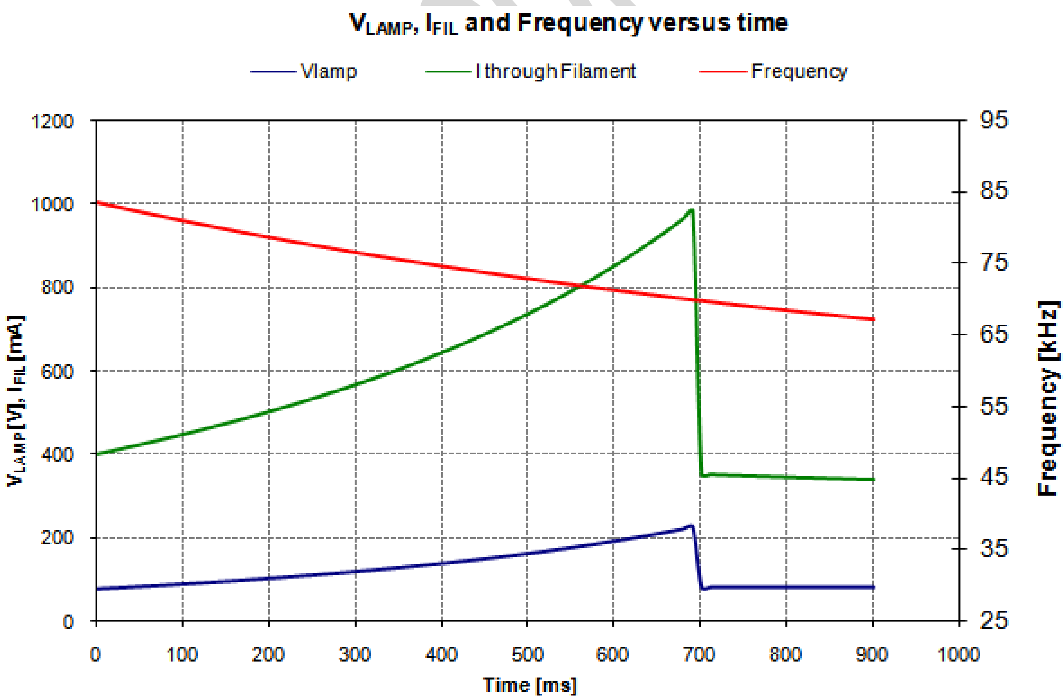


Fig 21. Burner voltage, filament current and frequency as a function of time during start-up

The tool will also show a graph of the calculated lamp power at f_{burn} ± 3 kHz, which will immediately warn the user in case a solution is on a steep slope of the power transfer curve of the resonant tank. This graph is shown in Fig 22.

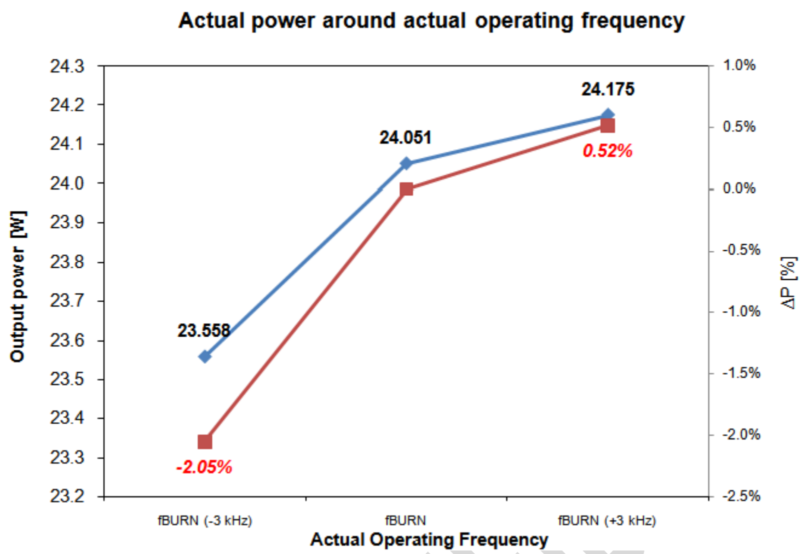


Fig 22. Calculated lamp power variation.

The power variation is shown in Watts as well as in percentage relative to the power at f_{burn}. The numerical output of the tool is shown in the Input / Output data fields (Fig 23).

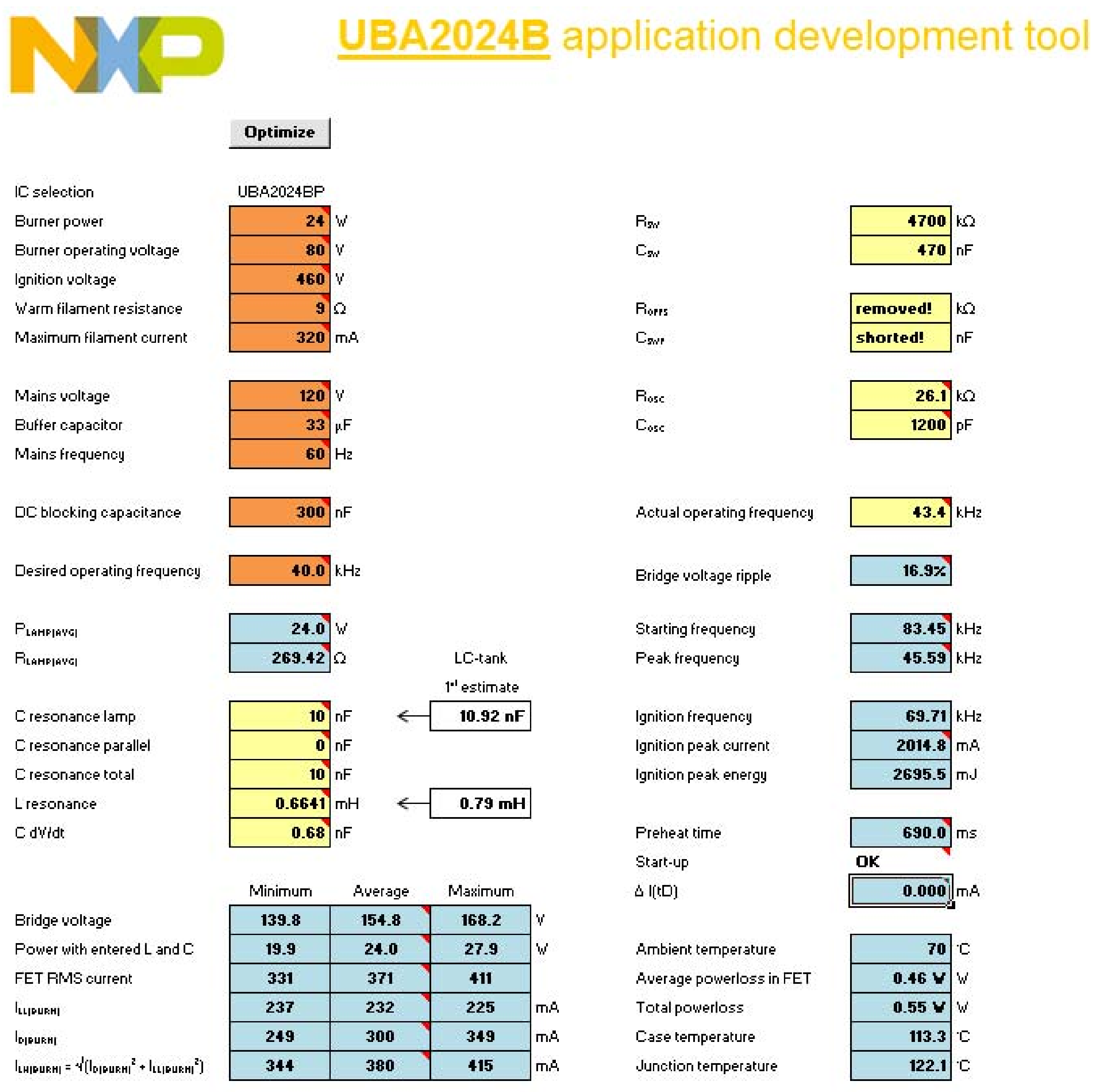


Fig 23. The Input / Output data fields

Next to the entered burner properties, the mains voltage and the wanted operating frequency also calculation results are shown here., which comprise of:

- Component values of the resonant tank (C_{rs}, C_{rp}, L_r and $C_{dV/dt}$)
- Actual operating frequency
- Average lamp power and resistance
- Various frequencies (f_{start}, $f_{ignition}$, f_{MIN}, f_{PEAK})
- The preheat time ($t_{PREHEAT}$)
- Start-up condition
- Coil current balance during the non-overlap time ($\Delta I_{(tD)}$)
- The advised resonant tank components which are calculated at the beginning of the algorithm ($C_{R(ADV)}$ (total capacitance of C_{rs}+ C_{rp}, not rounded to E12 values) and $L_{R(ADV)}$)
- Ignition peak current and energy through the coil
- Power dissipation and IC temperatures (case and junction)
- The advised oscillator resistance for a preheat time of 600 ms ($R_{OSC(ADV)}$)
- The voltage across the filaments (V_{FILAM})
- The lamp currents in the burn state of the application ($I_{LL(BURN)}$, $I_{D(BURN)}$, $I_{LH(BURN)}$)
- The minimum and maximum rectified mains voltage ($V_{BRIDGE(MIN)}$, $V_{BRIDGE(MAX)}$-D)

5.4.1 Coil

After finishing the calculation the tool also returns the most important coil requirements (example in Fig 24). These together with the inductance entered in Fig 23 and the operating temperature of the inductor is enough information to design a coil. Due to losses in the inductor the operating temperature of the inductor is higher than the lamp ambient temperature. When the coil is properly designed, the inductor temperature rise will be around 40 °C above the ambient temperature. In case a warm lamp is switched off and on again, the inductor should not saturate at this inductor temperature.

Ignition frequency	69.71	kHz
Ignition peak current	2014.8	mA
Ignition peak energy	2695.5	mJ

Fig 24. Coil design parameters

5.4.2 Thermal properties

In this section the estimated dissipated power and the estimated junction temperature in the IC is calculated. See Fig 25 for an example. When the maximum ambient temperature at which the lamp needs to operate is entered, the expected junction temperature is calculated. The junction temperature must not exceed 150 °C. If the junction temperature does exceed the 150 °C the expected operating life time of the IC is reduced significantly.

The maximum stress allowed during the ignition phase is 2500 mA (peak) on the UBA2024B at a case temperature of 25 °C (repetition rate is less than once per hour). The maximum stress period must not be longer than 1 second.

Ambient temperature	70	°C
Average powerloss in FET	0.42	W
Total powerloss	0.51	W
Case temperature	110.4	°C
Junction temperature	118.5	°C

Fig 25. Dissipated power and expected case and junction temperature in the IC

5.5 Choosing the other components

For the rectifier bridge a bridge cell or separate diodes like the 1N5062 can be used. The 1N4007 can also be used but these diodes not avalanche rugged.

For a lamp current ≥ 150 mA with $C_{dV/dt}$ = 220 pF and for a current ≥ 150 mA with $C_{dV/dt}$ = 100 pF the value of C_{vdd} and C_{fs} is 10 nF.

The advised half-bridge capacitors (C_{hb1} and C_{hb2}) are greater than 150nF when f_{out} = 40 kHz to 50 kHz and greater than 220 nF when f_{out} = 25 kHz to 30 kHz.

The resonance frequency of the input pi filter, consisting of L_{filt} and C_{hb} (CHB being the effective capacitor as seen on the HV pin of the IC, i.e. the series capacitance of C_{hb1} and C_{hb2}), has to be at least two times lower than the nominal output frequency.

Remark: Performance and lifetime cannot be guaranteed by using the values given in this chapter. The lamp and the UBA2024 performance strongly interact with each other and need to be qualified together as a combination.

5.6 Checking the tolerance sensitivity.

In this section the stability of the found result is verified with respect to mains fluctuations and component tolerances. After the tool has finished finding a solution how to dimension the application, the tool can be used how it reacts when for example the mains voltage changes ± 5 %, or when the resonant capacitors tolerances are taken into account.

Now an example is shown in Fig 26 when Vmains increases with 5 % from 120 V (AC) to 126 V (AC) what the effect is on the output power. Once the increased mains voltage is entered the actual burning frequency must be adapted such that Δ I(tD) equals zero again, so now the user optimizes for the operation on the edge of hard switching manually. In order to achieve the condition Δ I(tD) = 0 the frequency must be decreased. The field $P_{LAMP(AVG)}$ shows what the power is in this situation, and for the application that has just been designed this will result in an increase in power from 24 W to 26.376 W (+10 %).

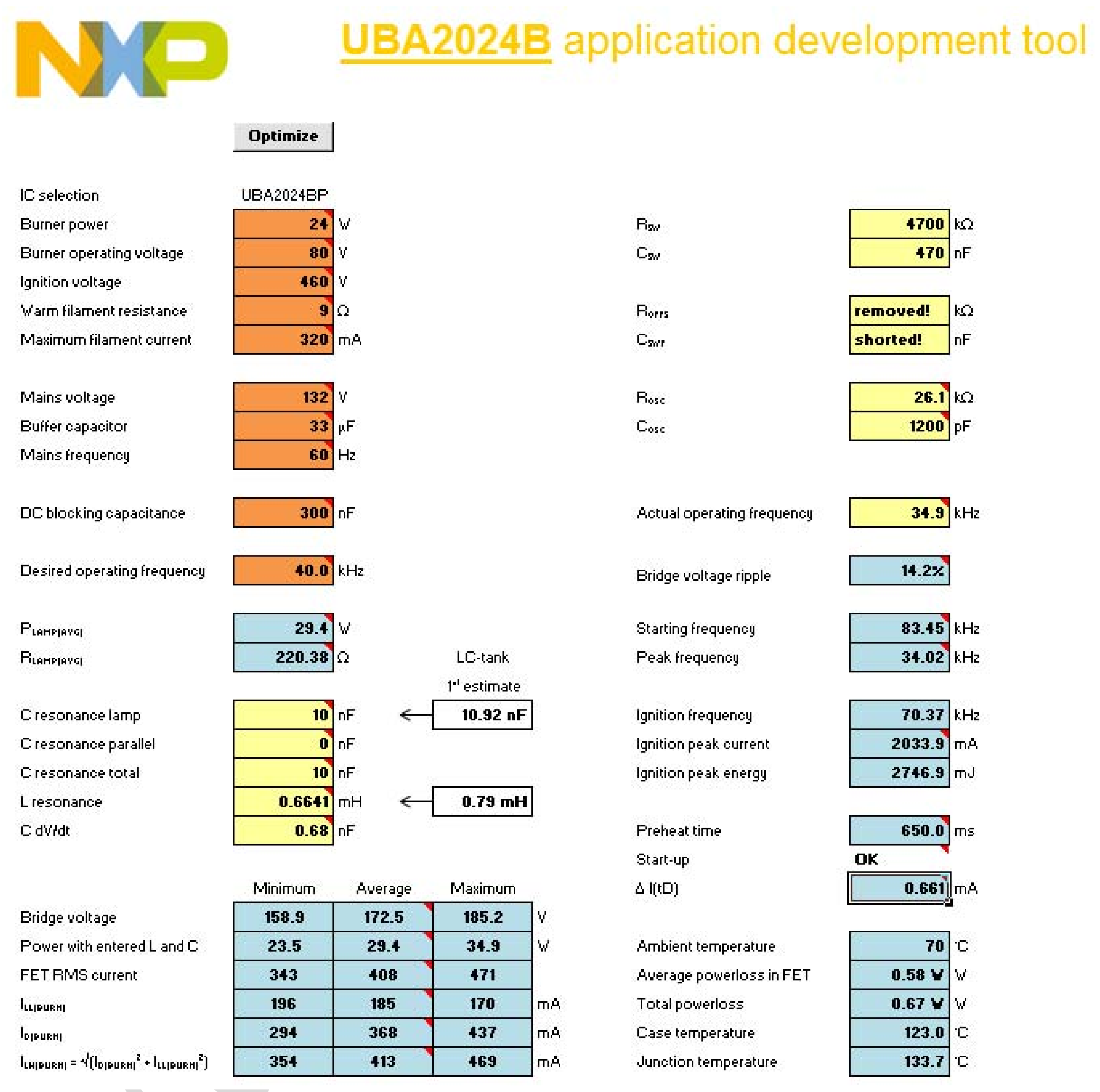


Fig 26. Sensitivity to mains variation (+ 10 %)

This procedure has also been carried out when tolerances of the resonant capacitors and inductor are taken into account. The result of this exercise is shown in Table 4.

Table 4. Calculated tolerance sensitivity results

Parameter (nominal)	V_{MAINS} = 120 V (AC) = 100 %		C_{rs} = 10 nF = 100 %		L = 0.66 mH = 100 %	
	V_{MAINS} (– 10 %)	V_{MAINS} (+ 10 %)	C_{rs} (– 5 %)	C_{rs} (+ 5 %)	L (– 10 %)	L (+ 10 %)
Frequency (43.4 kHz)	49.1 kHz	34.9 kHz	44.7 kHz	42.3 kHz	45.5 kHz	41.6 kHz
$P_{LAMP(AVG)}$ (24 W)	20.208 W	29.449 W	23.374 W	24.615 W	25.226 W	22.908 W

6. Building the application

6.1 Reference board

6.1.1 External lamp detection circuit

The NXP evaluation board has an additional lamp detection circuit which is not required in mass production applications like CFLi (see Fig 29). In this section the functioning of this detection circuit is described.

During start-up, in the preheat and the ignition phase, the voltage at the SW pin (pin 1) increases from 0 V to 1.32 V. At the same time the amplitude of the signal on the RC pin (pin 7) increases by the same amount. However, if the lamp is not ignited, because it is broken or missing, the sweep voltage will stay below the 3 V level or even drop to 0 V. The IC will not operate in Zero Voltage Switching mode (ZVS). Large currents run through the half-bridge causing the dissipation in the IC to exceed the maximum value. The half-bridge can only withstand the high dissipation until the junction temperature reaches 150 °C.

At start-up the RC oscillator starts with an amplitude of 2 V on pin RC (pin 8). The half-bridge frequency is now running at approximately 15 % above the nominal ignition frequency. When the burner is connected to the circuit the half-bridge operates in ZVS and the C_{sw} capacitor charges. R6, R7 and C12 create an average DC voltage of the oscillator voltage on pin RC, which is basically half the amplitude. That voltage is then fed to the base of Q2-2, which functions as a comparator.

At the same time that C_{sw} is charging, C11 is charged by R3 from VDD. This takes place with a time constant of (R3//R4) × C11. The charging stops when the voltage on C11 reaches 1.6 V. The voltage on C11 is fed to the emitter of Q2-2 to compare it with its base voltage.

Under normal conditions during start-up, when the lamp is connected, the average DC voltage from RC rises above 1.6 V at the end of the charging period for C11. The base emitter voltage of Q2-2 will stay reverse based and will not turn on. If, however, non-ZVS is detected in the switches of the half-bridge driver, because of an unconnected or broken lamp, the charging of C_{sw} stops and the voltage on C_{sw} drops to 0 V. The average DC voltage on the RC pin lowers to less than 1 V and Q2-2 starts to conduct.

Q2-2 drives the latching transistor Q1-1 and the fault condition is latched by the left diode of the double diode, D5. At the same time the right diode of D5 will stop the UBA2024B half-bridge oscillator. The latch can be reset by power cycling the mains voltage with less than 1 s delay (for the test circuit this depends on the discharge time of C11 and R4). The latch circuit is designed in such a way that it is not noise sensitive. However, it is better to keep it away from the large signal tracks.

Typically, the circuit triggers within 0.5 s from start-up when no lamp is connected. It also triggers when a lamp is removed while operating. When the protection has tripped, the dissipated power in the IC is about 0.6 W. The IC can dissipate this power continuously.

Ensure that there is some reaction time margin (at room temperature) when choosing C11. Also, consider voltage derating of MLCC capacitors when low voltage types are used. It is advisable to choose an X7R type or an X5R type of at least 10 V.

The protection circuit puts some additional capacitive loading (about 5 pF) on pin RC. For fixed frequency operation this can become significant for small values of C_{osc}. In this case the value of C_{osc} is compensated for this effect by lowering R_{osc} from 200 kΩ to 191 kΩ (E96 series), giving an operating frequency of 45.9 kHz instead of 43.3 kHz. When the circuit is used it is advisable to add the extra 5 pF to C_{osc} in Equation (3).

This additional capacitance can be neglected when the IC is working on the edge of hard switching, since a C_{osc} = 1.2 nF is recommended to improve the duty cycle of the half bridge output voltage.

Fig 27. Photo reference board UBA2024BP (DIL8)

Fig 28. Photo reference board UBA2024BT (SO14)

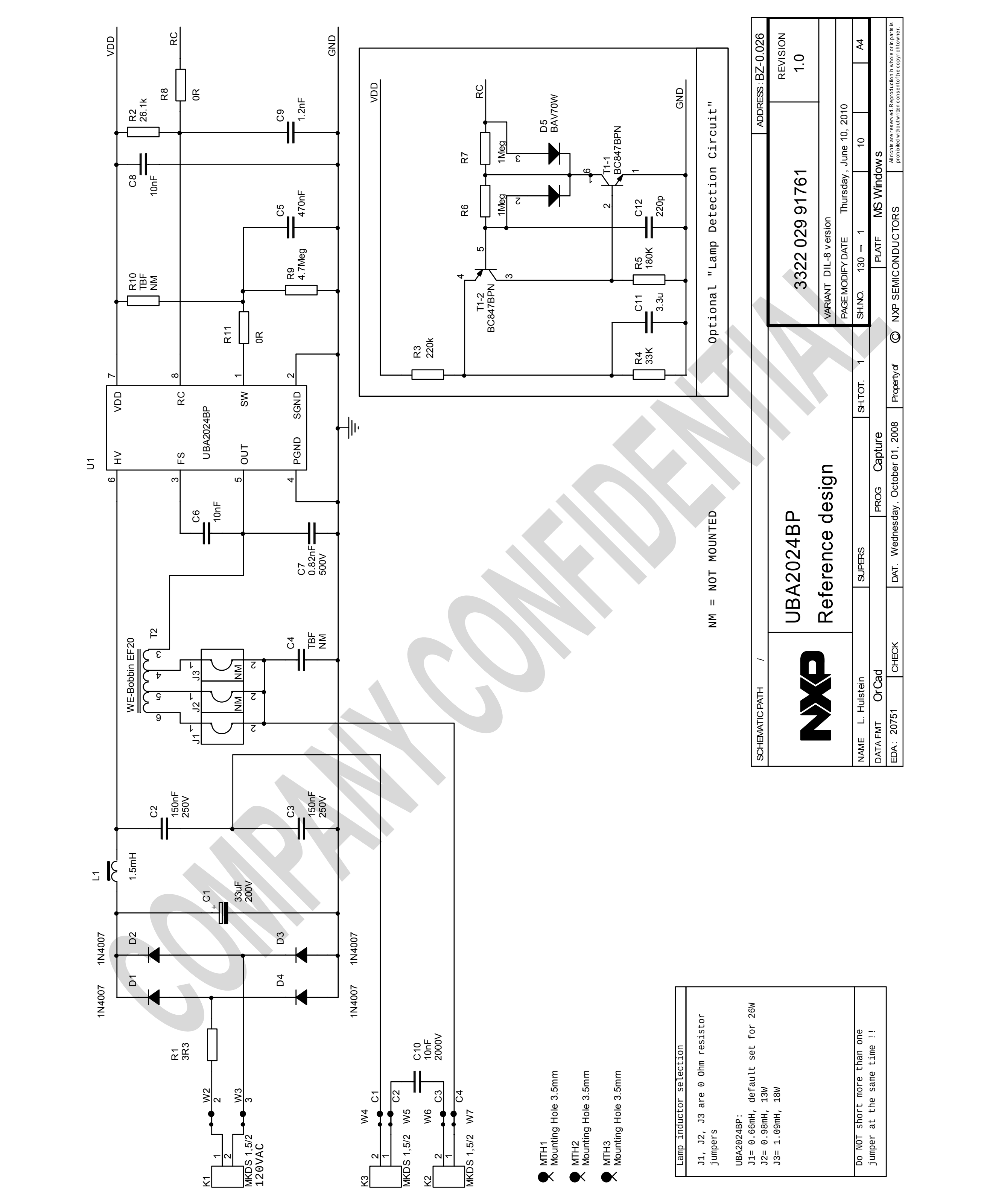


Fig 29. Circuit diagram of the UBA2024BP reference board with the optional lamp detection circuit

6.2 Bill of materials

In Table 5 the bill of materials is given for the application example with a PL-C 4P 26W lamp, including the external lamp detection circuit.

Table 5. Components used to build the application around the UBA2024BP for driving a PL-C 4P 26W CFL
This table applies to both the UBA2024BP and UBA2024BT referenceboard

Reference	Description	Remarks	Value
R1	Res., Fusible 3R3 / 5% / 2W NFR	Fusistor	3.3 Ω / 2 W
R2	Res. Thick film, 26K1 / 1% / 0W1 0603	Oscillator resistor	26.1 Ω / 0.1 W / 1 %
R3[4]	Res. Thick film, 220K / 5% / 0W1 0603		220 kΩ / 0.1 W
R4[4]	Res. Thick film, 33K / 5% / 0W1 0603		33 kΩ / 0.1 W
R5[4]	Res. Thick film, 180K / 5% / 0W1 0603		180 kΩ / 0.1W
R6, R7[4]	Res. Thick film, 1M / 5% / 0W1 0603		1 MΩ / 0.1W
R8, R11	Res. Thick film, 0R / 1% / 0W1 0603	Short	0 Ω
R9	Res. Thick film, 4M7 / 1% / 0W1 0603		4.7 MΩ / 0.1 W / 1 %
R10	n/a	Not mounted	n/a
C1	Cap, Al. El. 47uF / 20% / 200V KXG	High temperature electrolytic type	47 µF / 200 V
C2, C3	Cap. 150n / 10% / 250V DME		150 nF / 250 V
C4	n/a	Not mounted	n/a
C5	Cap. cer. 470n / 10% / 10V X5R 0603		470 nF / 10 V / 10 %
C6, C8	Cap. cer. 10n / 20% / 50V X7R 0603		10 nF / 50 V
C7	Cap. cer. 0.82n / 10% / 500V X7R 1206	dV/dt capacitor	0.82 nF / 500 V
C9	Cap. cer. 1n2 / 5% / 50V X7R 0603	Oscillator capacitor	1.2 nF / 50 V / 5 %
C10	Cap. 10n / 5% / 2KV MKP	Lamp capacitor	10 nF / 2 kV / 5 %
C11[4]	Cap. cer. 3u3 / 20% / 10V Y5V 0805		3.3 µF / 10 V
C12[4]	Cap. cer. 220p / 5% / 50V COG 0603		220 pF / 50 V / 5 %
D1, D2, D3, D4	Diode, Standard, 1KV, 1A	Mains rectifier diode	1N4007
D5[4]	Diode, small signal, dual, 70V 200mA	Double diode common cathode	BAV70W
L1	Ind. RF-Choke, 1m5H, 1R7, 0A43, 10%	Radial type	1.5 mH / 0.43 A
T1[4]	Tor., Dual, NPN/PNP 45V 100mA	PNP and NPN transistor in one	BC847BPN
T2	RF-Choke, T-H BOBBIN EF-20	E-20 core (select inductance with jumper)	0.66 mH / ~~0.88 mH~~ / ~~1 mH~~ (J1 in place)
U1[5]	UBA2024BP, UBA2024BT	CFL driver IC	UBA2024BP

[4] Component(s) needed for the optional lamp detection circuit

[5] There are two versions of the demo board available for a UBA2024BP and a UBA2024BT in a DIL8 and an SO14 package respectively

7. Layout considerations

The UBA2024B PCB layout has a considerable influence on the performance of the IC. Issues to be taken into account are:

- Coils with open magnetic circuits should not be placed opposite the IC (on the other side of the PCB). If an axial filter inductor is used for L_{filt} it should be placed in the same direction as the IC to minimize magnetic field pick-up.
- The oscillator pin (pin 7, RC) and the sweep pin (pin 8, SW) should be shielded from output/lamp by a ground track.
- Components on pins 7 and 8 should be placed as close to the IC as possible.
- Capacitors C_{vdd} and C_{fs} should be placed close to the IC.
- Mains input wires must not run parallel or near the half-bridge signal (pin 5, OUT) or near the output of the lamp inductor, bypassing the input filter.
- If the UBA2024BT is used, all SGND pins need to be soldered to a copper plane for effective heat transfer. This copper plane is underneath the IC and extends as much as possible on both sides of the IC. Fixing the IC to the board using thermal conductive glue also helps cooling the IC.

8. Quick measurements

Table 6 compares the the calculated values form the application development tool with the measured values. The measurements were carried out at 25 °C

Table 6. Measured values compared with the calculated values

	Lamp power [W]	f_{burn} [kHz]	$T_{preheat}$ [ms]	f_{start} [kHz]	f_{ign} [kHz]	I_{LL} [mA]	I_D [mA]	I_{LH} [mA]	$I_{ign(pk)}$ [mA]
Calculated	24	43.4	690	83.4	69.7	218	300	371	2010
Measured	23.4	51.5	560	81.7	70	279	279	404	1960

These components as calculated by the calculation spreadsheet for driving a PL-C 4P 26 W burner were used in the application.

L_{res} = 0.66 mH
C_{rs} = 10 nF
C_{rp} = not mounted
$C_{dV/dt}$ = 0.82 nF
R_{osc} = 26.1 kΩ
C_{osc} = 1200 pF.

9. Start-up waveforms

In the following two pictures the measured waveforms are shown of the lamp voltage, lamp current (Fig 30) and coil current (Fig 31) during preheat.

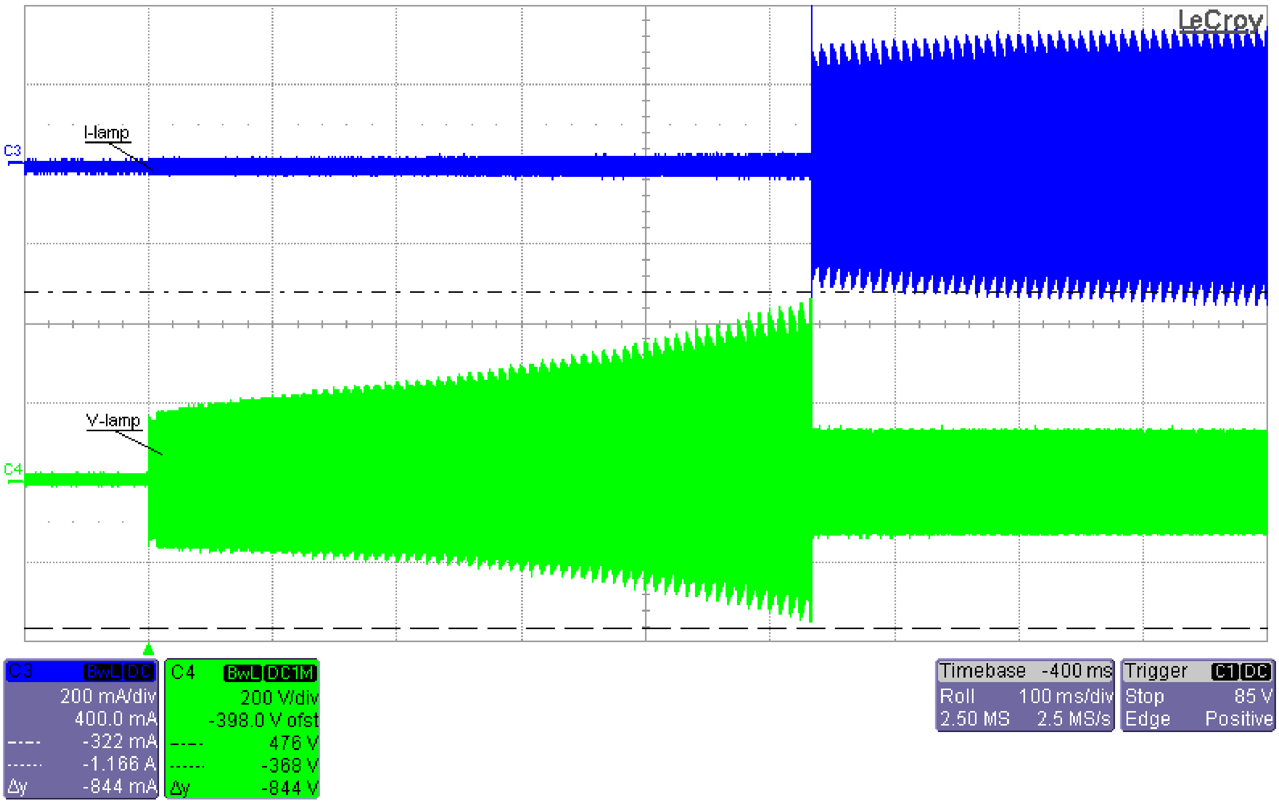


Fig 30. Startup waveforms showing lamp voltage and current, with L = 0.66 mH, C_{rs} = 10 nF, $C_{dV/dt}$ = 0.82 nF, C_{BUF} = 33 µF, R_{FUS} = 3.3 Ω [CSP205a.png]

Notice that the lamp ignites without glow. If lamp glow was present there would already be a lamp current before the lamp has ignited. This is not the case here. The ignition peak voltage amounts 470 V.

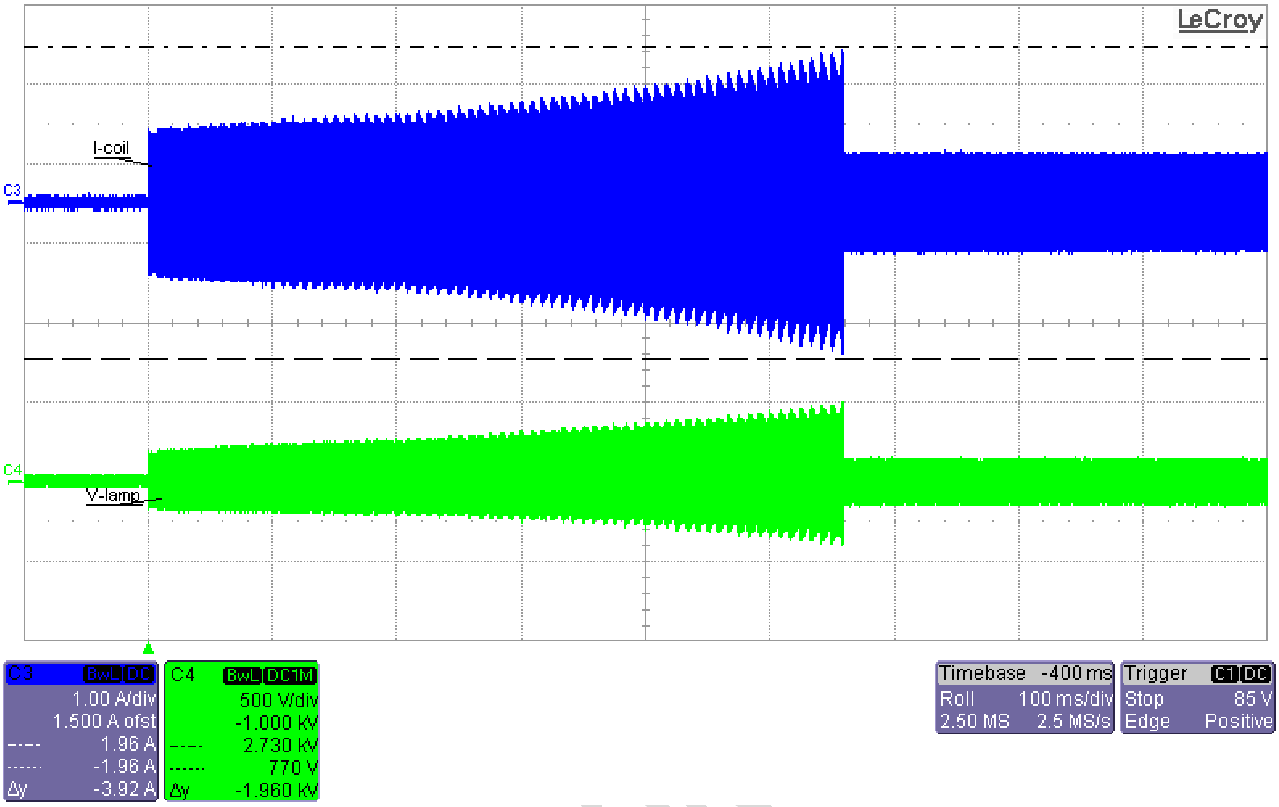


Fig 31. Startup waveforms showing lamp voltage and coil current [CSP206a.png]

The measured peak value of the coil current equals 1960 mA.

10. Steady state waveforms

Below the waveforms are shown 15 minutes after power on.

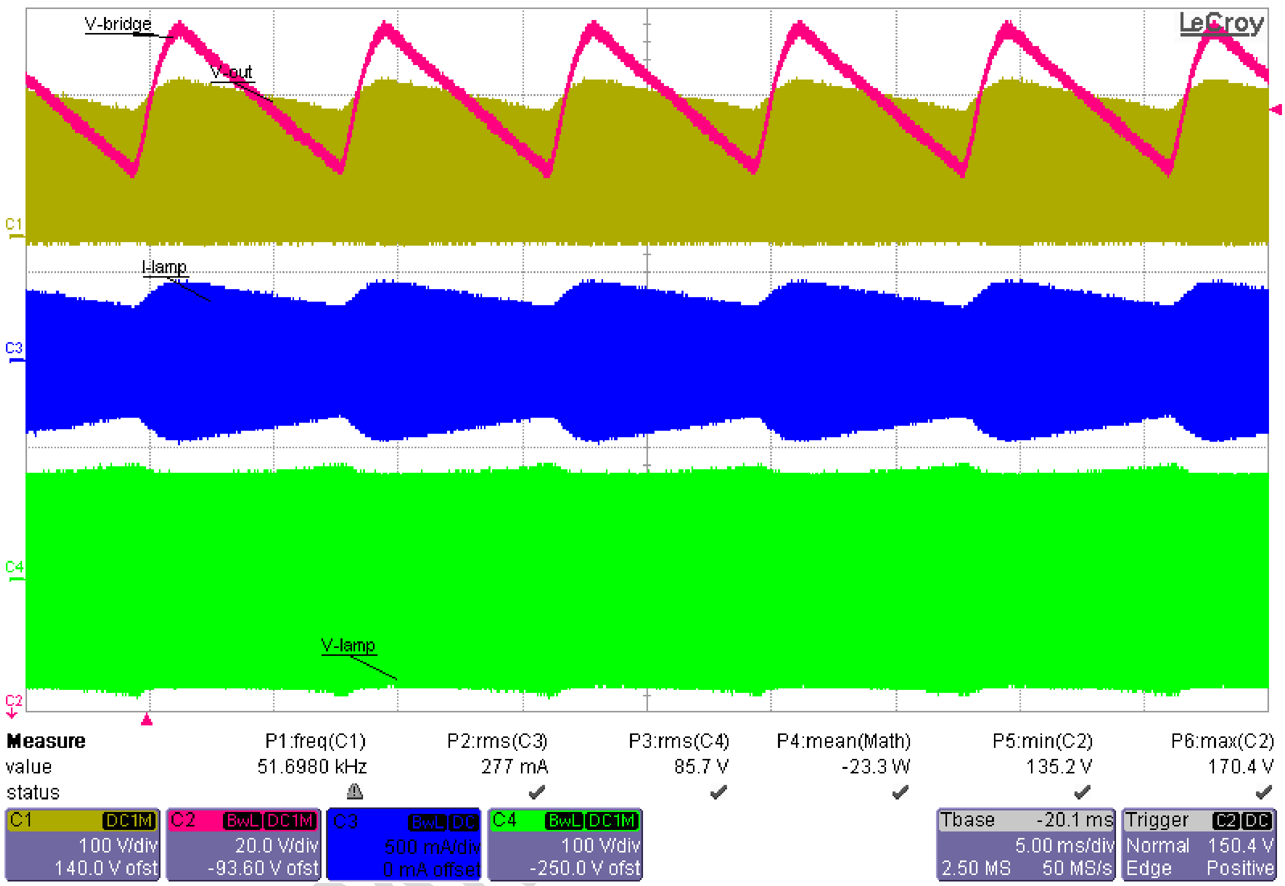


Fig 32. Steady state waveforms of the application at ambient temperature of 25 °C [CSP209.png]

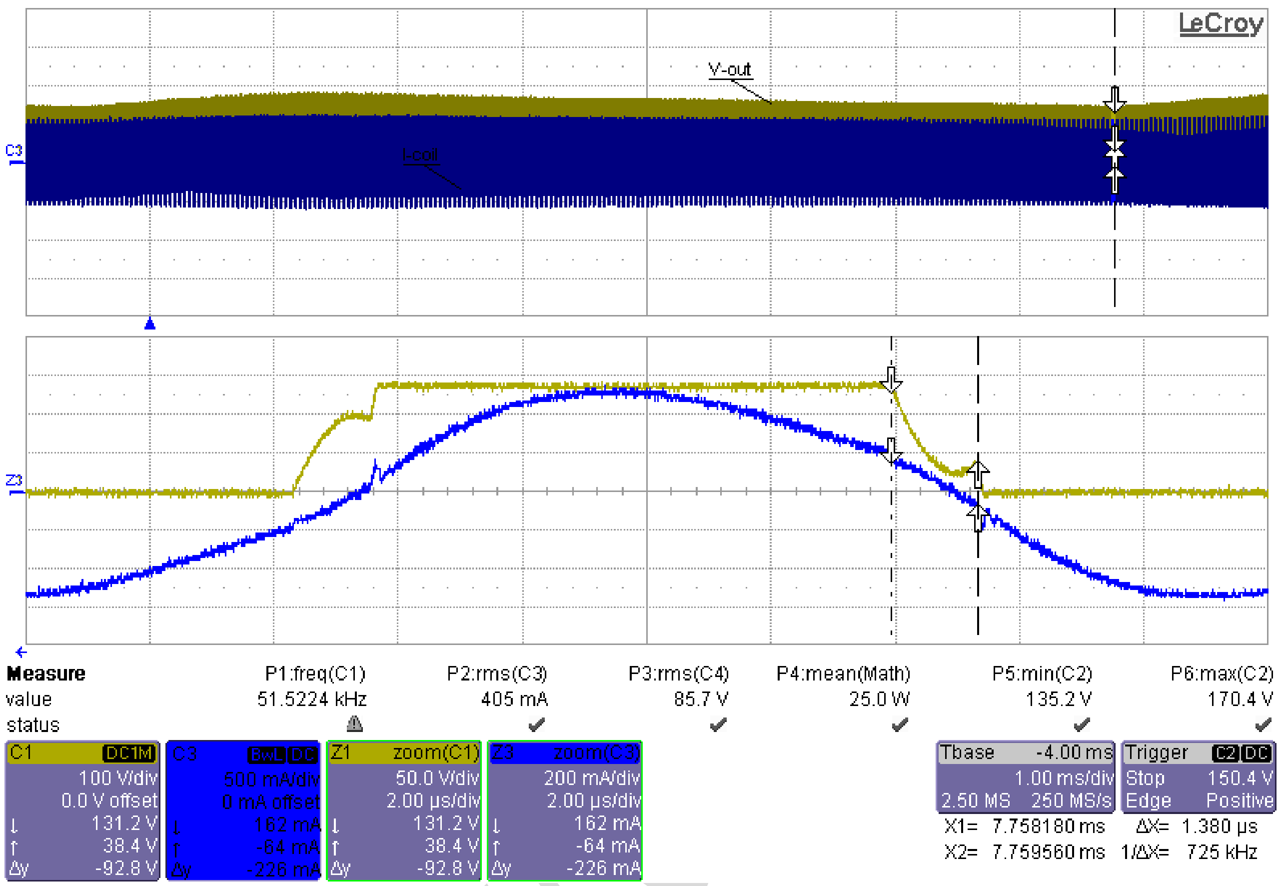


Fig 33. Steady state behavior, showing V-out and I-coil at $V_{bridge(MIN)}$ [CSP207.png]

In Fig 33 the measured coil current is shown during the dead time at the trailing edge of the half bridge voltage. Hard switching is seen here, while the frequency is controlled by zero crossing of the coil current. The shape of Vout has been changed such that the slopes are not so steep by increasing the calculated $C_{dV/dt}$ from 0.68 nF to 0.82 nF.

11. What if …

This section shows examples of practical problems like coil saturation and lamp glow.

11.1 Coil saturation

What happens when the coil goes into saturation during ignition in shown in Fig 34.

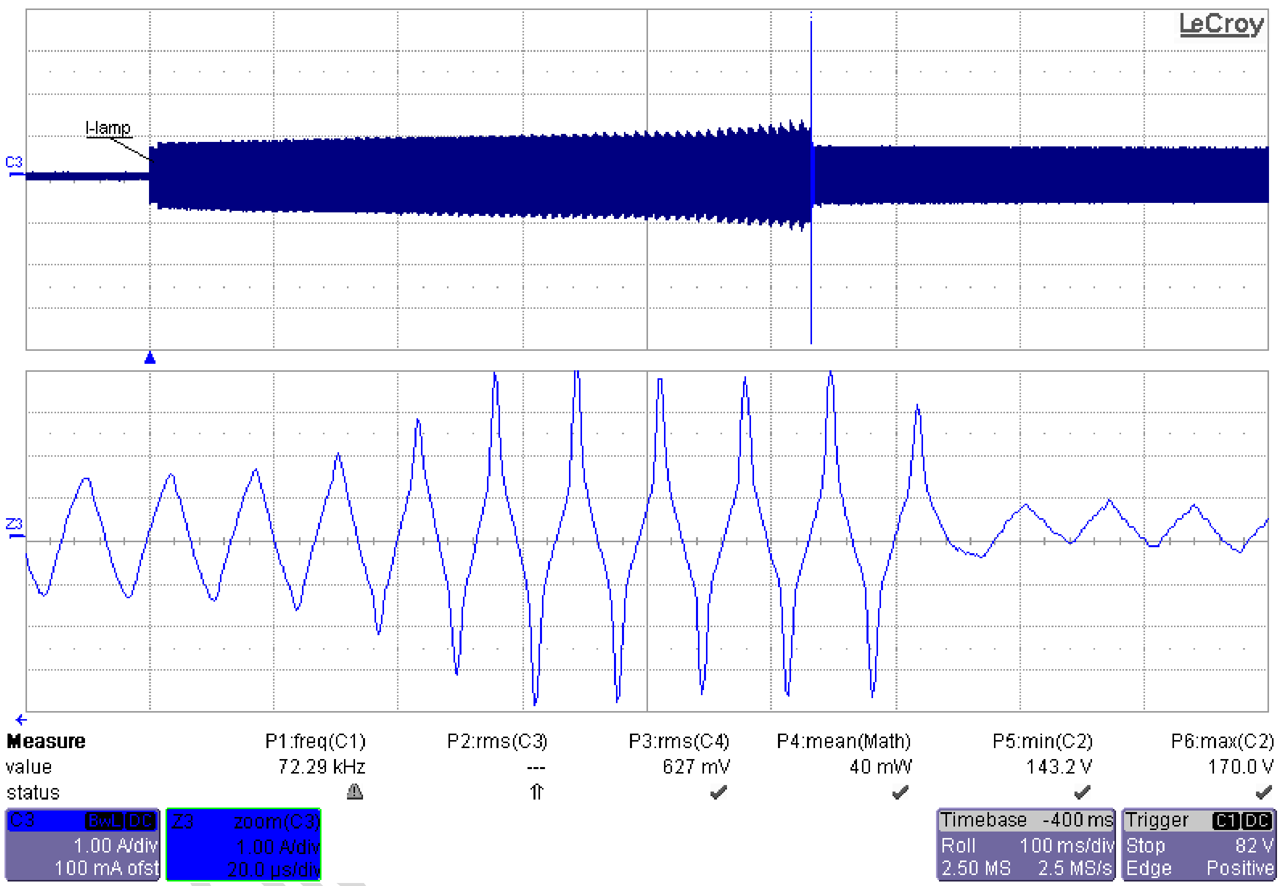


Fig 34. The coil current during ignition when the coil is saturated [CSP225.png]

In this case the coil current will show excessive peaks which in turn results in the integrated half bridge switches going into saturation and consequently damaging the IC.

11.2 Lamp glow

Lamp glow is mainly caused by improper preheating of the filaments.

Both a quasi-preheat that is too short and a voltage rise that is too slow increase the glow time of the lamp. This reduces the lifetime of the lamp. During the glow phase the lamp is ignited, but the filaments and the gas inside the lamp are not at their final operating temperature.

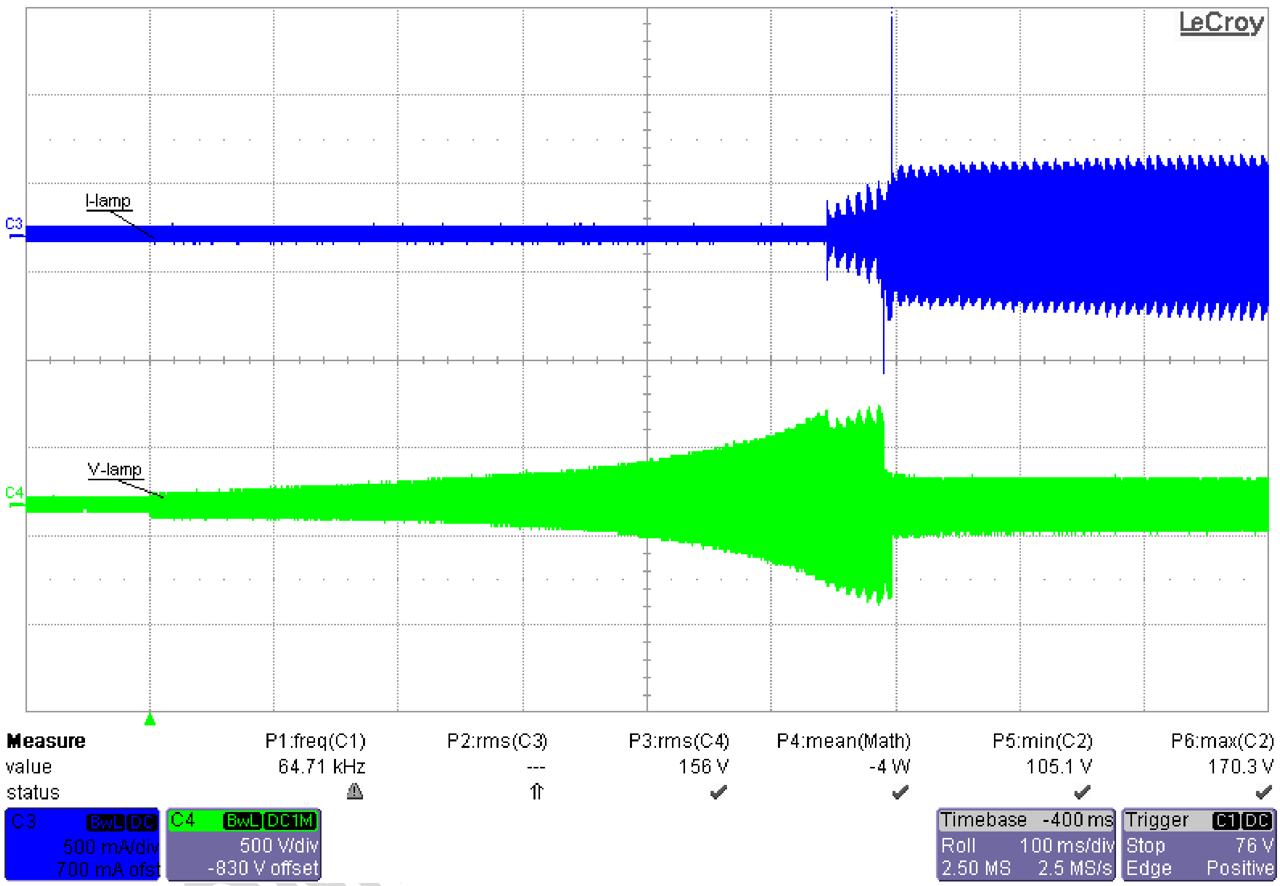


Fig 35. Lamp glow caused by improper preheating. [CSP221.png]

In Fig 35 it is clear that there is still a high voltage present over the lamp while at the same time lamp current is flowing. As soon as the filaments and gas inside the lamp have reached their normal operating temperature the voltage over the lamp will drop to its normal operating value.

This is the preheating method shown in Fig 16 referred to as Quasi preheat starting at 100 kHz.

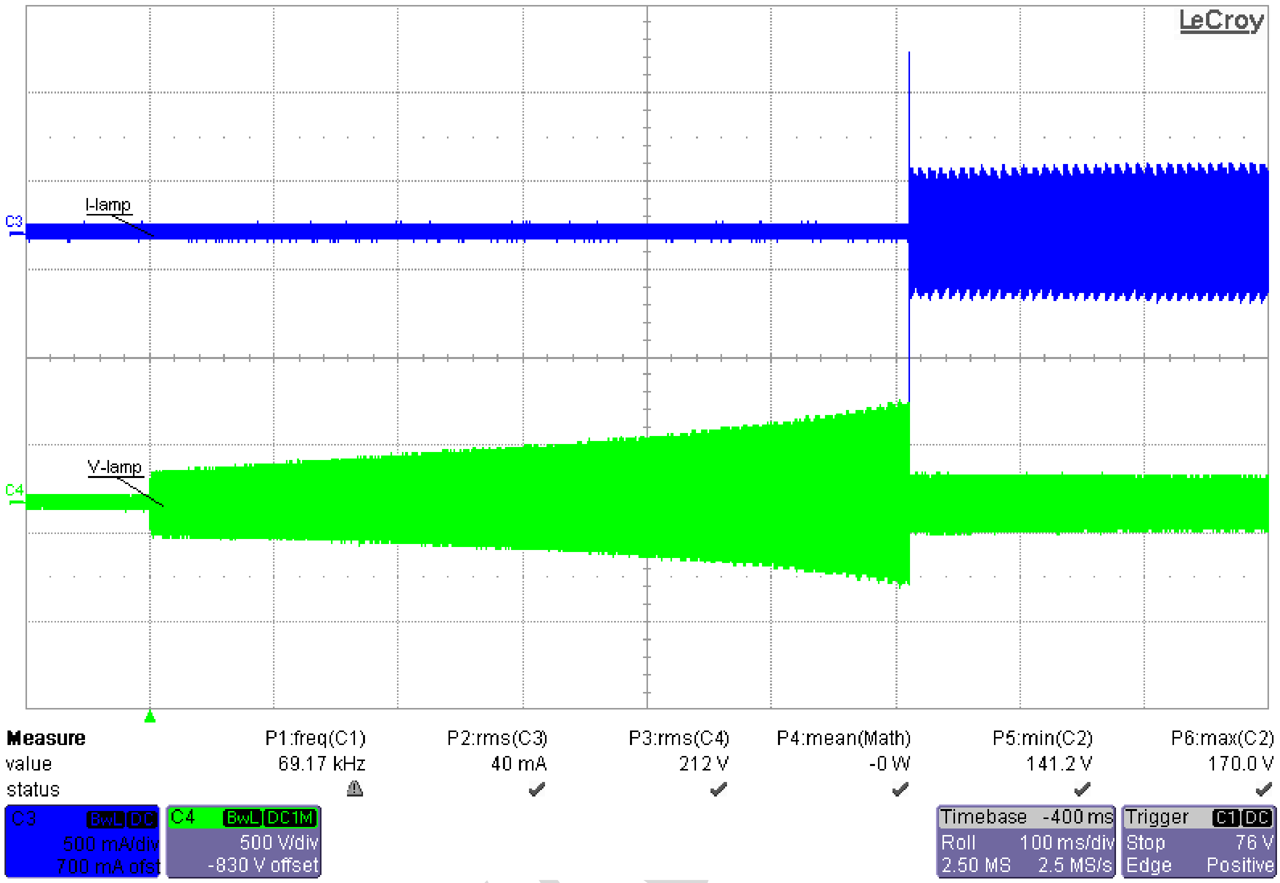


Fig 36. Proper ignition of the lamp due to proper preheating, without glow. [CSP222.png]

Fig 36 shows the ignition of a lamp that is preheated as shown in Fig 16 where preheating starts at the ignition frequency plus an additional 10 kHz. Note that there is no lamp glow present because the lamp's filaments had enough time to reach the right operating temperature. This method of preheating will increase the life time of the lamp and ensure passing a 10.000 (or more) times on/off test.

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