



Subject : **MHN-TD 70W Driver with UBA2030**

Pages: 39

Summary : This report describes a lamp driver concept intended for the MHL-TD 70W lamp (Metal Halide Lamp) used in general lighting applications. The circuit is a cascade of a buck converter and commutator. The latter is driven by the UBA2030. This concept is principally used to investigate the lamp behaviour, so it does not contain a preconditioner and is not EMC tested.

Keywords : MHN lamp; UBA2030; Buck Converter; Commutator; Igniter

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8. REVISION HISTORY 39

1. INTRODUCTION.

This report describes a lamp driver circuit intended for the MHL-TD 70W lamp (Metal Halide Lamp) belonging to the group of compact HID lamps (High Intensity Discharge lamps).

In contrast with TL and CFL lamps, HID lamps are high pressure discharge lamps. The presence of iodine will make starting more difficult compared to the low pressure discharge lamps mentioned before. An ignition voltage up to 25 kV can be needed to ignite an HID lamp in case of a hot re-strike. For more details about lamp behaviour, see chapter 2.

The HID lamps are applied in the automotive area and the general lighting area. Both application areas use the same lamp type but the operating requirements are rather different, so circuit topologies depend strongly on the application area. The system described in this report gives a driver concept for a general lighting application. This concept is principally used to investigate the lamp behaviour, so it does not contain a preconditioner and EMC is not tested and measured.

2. LAMP DATA.

The available lamp data is given in the next sub sections. Each sub section gives the characteristics of the lamp valid for the corresponding operating phase.

2.1 Ignition Phase

During the ignition phase, especially in case of a hot re-strike when the lamp is still warm from previous operation, a 25 kV voltage pulse should be supplied to the lamp with a fast rise-time. The total breakdown occurs within 10 ns. The voltage level is more important than the rise-time. The ignition voltage in case of a cold lamp amounts about 4-5 kV. So the ignition voltage increases with increasing lamp temperature.

2.2 Take-over phase

After ignition, the lamp voltage breaks down to a quarter of the nominal lamp voltage (20 V) and a conducting channel is produced. A take-over current should be applied in less than 1 ms preventing cooling down of the lamp. In most cases, the take-over current is derived from a capacitor which is loaded to a take-over voltage during the ignition phase and discharges to the lamp break-down voltage immediately after break down (20 V). The take-over current depends on the lamp resistance at break down. If the take-over current (or voltage) is too low, the lamp will extinguish and re-ignition is necessary to fire the lamp once more. Here again, the take-over voltage increases with increasing lamp temperature.

2.3 Run-up Phase

The lamp is characterised by the absence of a perceptible glow to arc transition. The lamp desires a relative high run-up current preventing extinguishing of the lamp. When the run-up current is too large, the electrode may be melting. The lamp voltage swells gradually to the nominal lamp voltage (85 V). The required run-up time is about 2 minutes and depends on the used luminaire.

2.4 Burn Phase

The lamp is designed to be driven with an AC current to avoid difference in electrode temperature which can cause acoustic resonance of the arc. However, re-ignition voltage pulses may be present when the lamp current is interrupted for a relative long time i.d. at the moment that the lamp current commutates. For example, the interruption time of a lamp driven with a sinusoidal voltage at 50 Hz will be relative large. The interruption time of a commutation circuit can be very short ($\leq 2 \mu\text{s}$)

Acoustic resonance occurs in the frequency domain: 10 kHz - 1 MHz, however some free-of-acoustic-resonance windows do exist. The commutating frequency of the full-bridge should be limited to the domain: 50 Hz - 10 kHz to avoid any risk on acoustic resonance. In practise, the commutating frequency range is limited to 100 Hz - 400 Hz.

The nominal lamp voltage (mean value 85 V) depends on the lamp type and the lamp aging effect, 65 V - 105 V. The differential resistance of the lamp is rather small and negative. To obtain a stable operating point, an impedance in series with the lamp is needed so that an ideal voltage source is

forbidden. The best lamp performance is achieved when the lamp is driven by a current source with power regulation. Lamp power control is important for the life-time of the lamp.

3. GENERAL CIRCUIT DESCRIPTION.

The complete circuit exists of two converters, namely:

- the buck converter which controls the lamp power and behaves like a current source.
- the full bridge converter which commutates the DC current from the buck converter into an AC current. Furthermore it supplies the igniter to generate ignition pulses.

The block diagram is given in figure 1 and the complete circuit diagram in figure 20 on page 37.

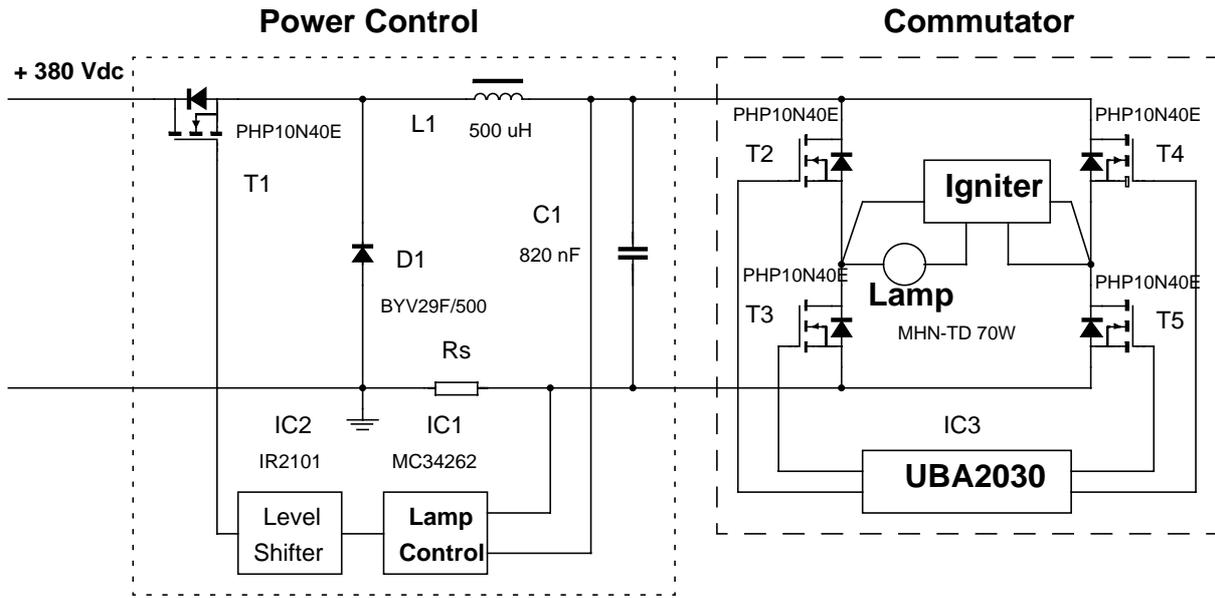


Fig.1 block diagram

3.1 Power Control

The power control is achieved by a buck converter in SOPS mode running at a relative high frequency of ± 70 kHz @ 85 V output voltage (lamp voltage). The SOPS mode operation is chosen to minimize the switch-on losses in the power mosfet. The lamp power is stabilised by a feedback signal formed by the addition of lamp voltage and lamp current to get a power controlled current source. The control IC1 is MC34262 which normally is used as power factor correction IC. The lamp power as function of lamp voltage is parabolical. The top of the parabola, the nominal lamp power P_n and nominal lamp voltage U_n , are set to $(P_n, U_n) = (70 \text{ W}, 85 \text{ V})$. The maximum lamp current is limited to 1 A, so the parabolical relation is not valid for lamp currents smaller than 1 A.

The maximum output voltage of the buck converter would be $2 \cdot U_n$ (170 V), due to the parabolical function, see figure 2 on page 13. So an ignition voltage control circuit is added to replace the power control by a voltage control when the voltage is larger than 140 V and forces the buck converter to act like a voltage source of 340 V. Now, the output voltage range is $>0 - 340 \text{ V}$, where the 340 V is used for the igniter to generate ignition pulses.

Immediately after break-down, the output voltage breaks down to zero and the buck converter is not able to operate due to the missing restart pulses. A Watch-dog Timer function is added to keep the buck converter running, see section 4.3.

Finally, a level-shift IC2 (IR2101) is used to drive the power mosfet of the buck converter.

3.2 Commutator & Igniter

The commutation is achieved by a full bridge converter driven by the UBA2030 with a commutation frequency of ± 100 Hz.

The igniter is connected between the two mid-points of the full bridge and generates ignition pulses of 4 kV peak with a duration of 100 μ s and a repetition frequency of 100 Hz (commutator frequency). The breakdown device is a sidac (break-over diode) with a typical breakdown voltage of 340 V.

4. POWER CONTROL.

4.1 Current Sensing

The current sensing block measures the buck converter coil current by three in parallel connected resistors R_1 , R_2 and R_3 called R_S (1.57Ω) followed by a voltage divider network R_4 and R_5 (both 100Ω) to obtain information about the actual peak current in the buck coil, see figure 20 on page 37. The peak current is controlled by a feedback signal at pin 1 of the IC1. The maximum peak current is set by an internal voltage limiter inside the IC, named $U_{th(max)} = 1.5 \text{ V}$. So the maximum lamp current $I_{la,max}$ is set to 1 A, see equation 1. The complete derivation is given in equation 13 on page 29.

$$I_{la,max} = \frac{k}{2} \cdot \frac{U_{th,max}}{R_S} \dots \text{with} \dots k = 1 + \frac{R_5}{R_4} \quad (1)$$

Data: $U_{th,max} = 1.5 \text{ V}$, $R_S = 1.57 \Omega$, $R_5 = 100 \Omega$ and $R_4 = 100 \Omega$.

Calculations: $I_{la,max} = 0.96 \text{ A}$.

Another sensing signal is evident for proper operation of the buck converter namely, the zero current detection at pin 5 of IC1. This signal senses indirectly the inductor current by monitoring the voltage across the auxiliary winding on the buck coil. When it falls below the level of 1.4 V , it initiates the turn-on of the buck mosfet.

During the run-up phase, the output voltage of the buck converter oscillates at every commutation for some milliseconds down to zero, see section 5.1 on page 19. The zero current detection circuit is not able to generate restart pulses because the secondary voltage stays for a relative long time lower than 1.4 V . To overcome this problem a Watch-dog Timer is added to keep the buck converter going on, see section 4.3 on page 13.

4.2 Power Control Management

The feedback signal should be a multiplication of the lamp voltage and -current but for simplicity an addition is used to obtain power control. But the floating lamp makes sensing difficult, so the lamp voltage is sensed across the buck converter capacitor C_1 and the lamp current is represented by the voltage across R_S (1.57Ω). The indirect lamp current sensing is valid because the average buck coil current is equal to the lamp current (SOPS mode) and the average voltage across C_1 is equal to the lamp voltage. Addition of both lamp voltage and -current results in a parabolical relation between lamp power and lamp voltage. Lamp power variation due to lamp voltage variation are the smallest when the nominal lamp power is set to the top of the parabola.

The lamp voltage is measured by R_U (R_6 in series with $R_7 \rightarrow 550 \text{ k}\Omega$) and the lamp current by R_i ($8.2 \text{ k}\Omega$) which is the sum of R_{i1} (R_{10} in parallel with $R_{11} \rightarrow 4 \text{ k}\Omega$) and R_{i2} (R_8 in parallel with $R_9 \rightarrow 4.2 \text{ k}\Omega$). Equation 2 gives the relation between the nominal lamp power P_n , nominal lamp voltage U_n , the internal reference voltage U_{ref} and sensing resistors R_U , R_i and R_S .

$$\alpha_n = \frac{R_U}{R_i} = 2 \cdot \frac{U_n}{U_{ref}} - 1 \quad R_S = \frac{U_n^2}{P_n \cdot \alpha_n} \quad (2)$$

Substituting the lamp data and IC data gives the values for the used components.

Data: $U_{ref} = 2.5 \text{ V}$, $P_n = 70 \text{ W}$ and $U_n = 85 \text{ V}$.

Calculations: $\alpha_n = 67$, $R_s = 1.54 \text{ } \Omega$.

R_s is formed by three PR02 resistors of $4.7 \text{ } \Omega$ connected in parallel, so the actual value of the sense resistor R_s is $1.57 \text{ } \Omega$.

The complete derivation of $P_{Ia}(U_{Ia})$ and $I_{Ia}(U_{Ia})$ is given in equation 14 through 20 in appendix A.1.2 on page 29 and the end result is given in equation 3 and 4. Figure 2 gives the corresponding curves.

Equation 3 gives the relation between $P_{Ia}(U_{Ia})$ and $I_{Ia}(U_{Ia})$. Two regions are distinguished, namely:

- region 1, the region where $P_{Ia}(U_{Ia})$ is linear and $I_{Ia}(U_{Ia})$ constant.
- region 2, the region where $P_{Ia}(U_{Ia})$ is parabolical and $I_{Ia}(U_{Ia})$ linear.

Both are caused by the lamp current limitation $I_{Ia,max}$ described in section 4.1.

Equation 4 gives the boundary values between region 1 and region 2, namely:

- $U_{Ia,min}$, the minimum lamp voltage where the parabolical relation of $P_{Ia}(U_{Ia})$ is valid.
- $U_{Ia,max}$, the maximum lamp voltage due to the parabolical relation of $P_{Ia}(U_{Ia})$.
- P_{br} , the lamp power at the breakpoint. Note that $P_{br} \leq P_n$.

$$\begin{cases} I_{Ia} = \frac{P_n}{U_n} \cdot \left(2 - \frac{U_{Ia}}{U_n}\right) & U_{Ia} \geq U_{Ia,min} \\ I_{Ia} = I_{Ia,max} & U_{Ia} \leq U_{Ia,min} \end{cases} \quad (3)$$

$$\begin{cases} P_{Ia} = P_n \cdot \frac{U_{Ia}}{U_n} \cdot \left(2 - \frac{U_{Ia}}{U_n}\right) & U_{Ia} \geq U_{Ia,min} \\ P_{Ia} = U_{Ia} \cdot I_{Ia,max} & U_{Ia} \leq U_{Ia,min} \end{cases}$$

$$U_{Ia,min} = U_n \cdot \left(2 - \frac{I_{Ia,max} \cdot U_n}{P_n}\right) \quad P_{br} = I_{Ia,max} \cdot U_n \cdot \left(2 - \frac{I_{Ia,max} \cdot U_n}{P_n}\right) \quad (4)$$

From calculations in section 4.1, we know that the maximum lamp current $I_{Ia,max} = 0.96 \text{ A}$. Substitution of $I_{Ia,max}$ in equation 4 gives a value for $U_{Ia,min}$ and P_{br} .

Data: $I_{Ia,max} = 0.96 \text{ A}$, $P_n = 70 \text{ W}$ and $U_n = 85 \text{ V}$.

Calculations: $U_{Ia,min} = 71 \text{ V}$ and $P_{br} = 68 \text{ W}$.

Note that the maximum output voltage is limited to $2 \cdot U_n = 170 \text{ V}$ due to the position of the parabola. This means that the igniter can not generate ignition pulses. An Ignition Voltage Control circuit solves this problem, see section 4.4.

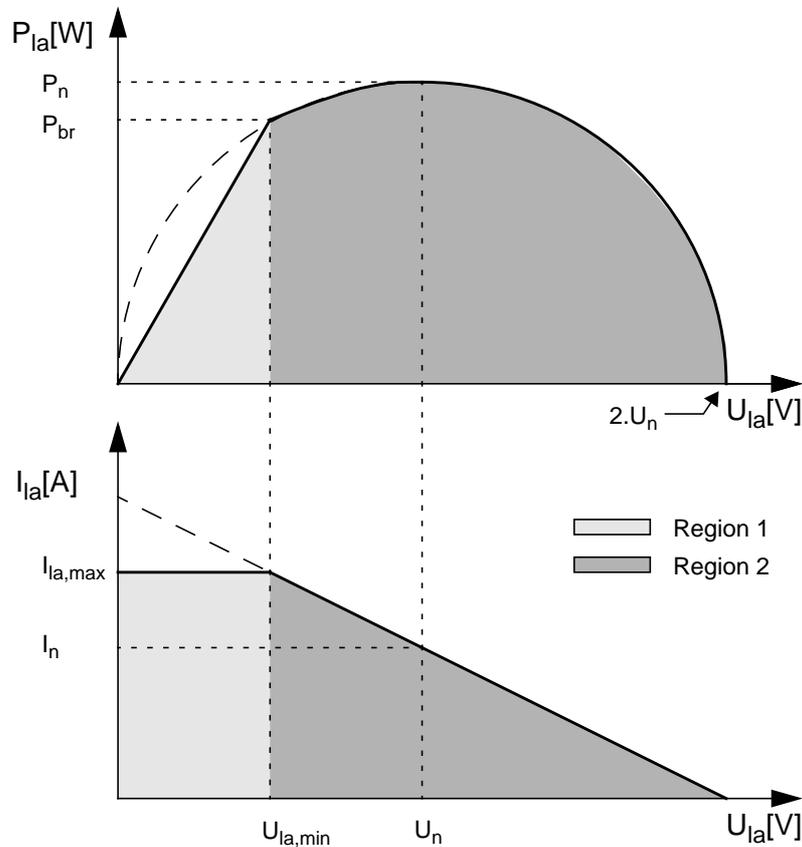


Fig.2 $P_{la}(U_{la})$ and $I_{la}(U_{la})$

4.3 Watch-dog Timer

The buck converter output voltage range is rather large >0 - 340 V. After ignition, the output voltage breaks down to zero and the zero current detection signal, which initiates the turn-on of the buck mosfet, is not able to generate pulses anymore see section 4.1. So a watch-dog timer is applied to generate a restart pulse when IC1 does not generate pulses for a time longer than t_{wd} seconds. The delay time t_{wd} is given by equation 5 derived in equation 21 on page 32 where E is the low voltage supply, $\tau_{wd} = R12 \cdot C5$ and $U_{ref(s)}$ the system reference voltage.

$$t_{wd} = \tau_{wd} \cdot \ln\left(\frac{E}{E - U_{ref(s)}}\right) \quad (5)$$

Data: $E = 15$ V, $U_{ref(s)} = 5$ V, $R12 = 12$ k Ω and $C5 = 10$ nF. Calculations: $\tau_{wd} = 120$ μ s and $t_{wd} = 49$ μ s. The watch-dog timer key components are R12, C5, TR6 and IC4-1.

4.4 Ignition Voltage Control

Due to power control method described in section 4.2 on page 11, the maximum output voltage of the buck converter is limited to $2 \cdot U_n = 170 \text{ V}$. The Igniter circuit needs 360 V, so an Ignition Voltage Control circuit is added to solve this problem. The Ignition Voltage Control circuit modifies the feedback loop twofold. The current sensing information across R_s is inhibited so as a result the buck converter behaves like a voltage source. Simultaneously, the feedback resistor divider ratio is adapted by the open collector output of IC4-2 to generate the correct voltage on C1 needed for ignition $U_{C1} = U_{C1,ign}$. The Ignition Voltage Control circuit is activated when the output voltage of the buck converter U_{C1} is larger than the trip-level U_{trip} . U_{trip} is fixed by R20, R21, R22 and $U_{ref(s)}$, see equation 6.

$$\begin{aligned} U_{C1,ign} &= \left(1 + \frac{R_u}{R_{i1}}\right) \cdot U_{ref} \\ U_{trip} &= \left(1 + \frac{R_{20} + R_{21}}{R_{22}}\right) \cdot U_{ref(s)} \end{aligned} \quad (6)$$

Data: $U_{ref} = 2.5 \text{ V}$, R_u (R6 in series with R7) = 550 k Ω , R_{i1} (R10 in parallel with R11) = 4 k Ω .

$U_{ref(s)} = 5 \text{ V}$, R20 = 150 k Ω , R21 = 120 k Ω and R22 = 10 k Ω .

Calculation: $U_{C1,ign} = 346 \text{ V}$.

$U_{trip} = 140 \text{ V}$.

Note: $U_{C1,ign}$ is an averaged value. Just after ignition, U_{C1} is smaller than 360 V due to the large ignition currents. So, just before ignition, U_{C1} must be larger than 360 V to retain the averaged value of 346 V.

4.5 Level Shift Driver

The Level Shift Driver function is necessary to drive the floating power N-mos TR1. The most important characteristic of the driver IC2 is that the propagation delay between input and output signal is small and the maximum gate drive current is large. Both take care of fast turn-on and turn-off switching of the power mosfet to minimize switching losses. However, the gate drive current of the applied Level Shift Driver is rather small so that the mosfet is turned on with a small delay and optimal switching is not achieved.

The driver sink current and the total parasitic capacitance parallel with the mosfet determine the switch off losses of the mosfet, see section 4.6.2.

The floating power supply is formed by bootstrap diode D3 and bootstrap capacitor C6 which is charged to 15 V when the buck diode D1 is conducting. During ignition, the current through the buck diode is zero and the voltage on the source of the power mosfet stays larger than 15 V so that the bootstrap diode is not loaded anymore and the mosfet can not be turned on. Now, an additional winding on the buck coil L1 provides the floating power supply. The polarity is in phase with the buck output voltage.

4.6 Power Components

4.6.1 Buck Coil

The buck coil L1 is an EF32 -core with 3C85 as core material and 63 turns 10*0.2 mm litzwire. Two auxiliary windings of 6 and 3 turns are used for successively zero current detection and additional floating supply for the level shift driver IC1, see respectively section 4.1 and 4.5. The value of the inductor is $\pm 490 \mu\text{H}$ and the wire resistance $240 \text{ m}\Omega$.

The operating frequency f_{op} depends on the value of the buck coil as you can see in equation 7. Equation 7 is derived from equation 22 through 24 on page 33 and is only valid when $T_{\text{osc}} \ll 2/f_{\text{op}} \cdot T_{\text{osc}}$ is the freewheel oscillation period time caused by the buck coil L1 and C_{par} the total parasitic capacitance from source to ground, $C_{\text{par}} = C_{\text{gd}} + C_{\text{ds}} + C_{\text{L1}} + C_{\text{D1}}$. The oscillation starts when the current through the buck diode D1 drops to zero and stops conducting.

$$f_{\text{op}} = \frac{U_{\text{Ia}} \cdot U_{\text{n}}}{2 \cdot L \cdot P_{\text{n}}} \cdot \frac{1 - \frac{U_{\text{Ia}}}{U_{\text{in}}}}{2 - \frac{U_{\text{Ia}}}{U_{\text{n}}}} \quad \left(\begin{array}{l} U_{\text{Ia}} \geq U_{\text{Ia, min}} \\ I_{\text{Ia}} \leq I_{\text{Ia, max}} \end{array} \right) \quad (7)$$

4.6.2 Power Mosfet

The power mosfet TR1 is of the Philips type PHP10N40E (BUK457/400) with a typical $R_{\text{ds(on)}} = 0.55 \Omega$. The conduction losses $P_{\text{m,c}}$ are given by equation 8. The complete derivation is given in equation 26 on page 33.

$$P_{\text{m,c}} = \frac{4}{3} \cdot R_{\text{ds(on)}} \cdot \frac{P_{\text{Ia}}^2}{U_{\text{Ia}} \cdot U_{\text{in}}} \quad (8)$$

Data: $P_{\text{n}} = 70 \text{ W}$, $U_{\text{n}} = 85 \text{ V}$, $U_{\text{in}} = 380 \text{ V}$ and $R_{\text{ds(on)}} = 0.55 \Omega$.

Calculations: $P_{\text{m,c}} = 111 \text{ mW}$.

The switch-on losses $P_{\text{m,son}}$ are given by equation 9. The complete derivation is given in equation 27 and 28 on page 34. The operating frequency f_{op} of the buck converter is determined by the value of the buck coil L1, the input voltage U_{in} , the output voltage U_{Ia} , the output power P_{Ia} and the total parasitic capacitance at the source of the mosfet C_{par} . C_{par} and L1 determine the freewheel oscillation period time T_{osc} when the current through the buck diode drops to zero.

$$P_{\text{m,son}} = f_{\text{op}} \cdot \frac{1}{2} \cdot C_{\text{par}} \cdot (U_{\text{in}} - 2 \cdot U_{\text{Ia}})^2$$

$$C_{\text{par}} = \left(\frac{T_{\text{osc}}}{2 \cdot \pi} \right)^2 \cdot \frac{1}{L} \quad (9)$$

$$f_{\text{op}} = \frac{U_{\text{Ia}}^2}{2 \cdot L \cdot P_{\text{Ia}}} \cdot \left(1 - \frac{U_{\text{Ia}}}{U_{\text{in}}} \right)$$

Data: $P_n = 70 \text{ W}$, $U_n = 85 \text{ V}$, $U_{in} = 380 \text{ V}$, $T_{osc} = 1.5 \mu\text{s}$, $L = 500 \mu\text{H}$

Calculations: $f_{op} = 80 \text{ kHz}$, $C_{par} = 115 \text{ pF}$, $P_{m,son} = 200 \text{ mW}$

The switch-off losses are described in equation 29 through 31 on page 36 and the end result is given in equation 10. Equation 10 gives an approximation of the real switch-off losses where some non-linear quantities like C_{ds} , C_{dg} and g_m are supposed to be linear. So, equation 10 gives merely the switch-off behaviour and not the quantitative switch-off losses, with $C_p = C_{ds} + C_{L1} + C_{D1}$.

$$P_{m,soff} = f \cdot \frac{1}{2} \cdot C_{gd} \cdot U_{in}^2 \cdot \left(\frac{\hat{I}_L}{I_g} - k \right) \quad (10)$$

$$k = 1 + \frac{C_p}{C_{gd}}$$

From equation 10, we can see that the switch-off losses can be reduced by increasing the gate sink current I_g or increasing the capacitance C_p . The disadvantage of increasing C_p is that the switch-on losses will increase due to the increase of capacitance C_{par} in equation 9, $C_{par} = C_{gd} + C_p$.

Calculations are not given because the difference between the approximated relation in equation 10 and reality is too large. From measurements it could be derived that an extra capacitance C_{ext} of 270 pF in parallel with diode D1 reduces the overall switching losses substantially, see figure 9 and 10 on page 23. The figures 11 through 14 give a more detailed picture of the switch-on and switch-off behaviour depending on C_{ext} .

4.6.3 Diode

The buck diode D1 is of the Philips type BYV29F/500V. The diode conduction losses $P_{d,c}$ will be a first order approximation described in equation 32 through 34 on page 36. The end result, given in equation 11, needs two input arguments; U_f as diode forward voltage and R_{dio} as diode resistance. Both values are measured on a curve tracer.

$$P_{d,c} = U_f \cdot \frac{P_{Ia}}{U_{Ia}} + R_{dio} \cdot \frac{4}{3} \cdot \left(\frac{U_{in}}{U_{Ia}} - 1 \right) \cdot \frac{P_{Ia}^2}{U_{Ia} \cdot U_{in}} \quad (11)$$

Data: $P_n = 70 \text{ W}$, $U_n = 85 \text{ V}$, $U_{in} = 380 \text{ V}$, $U_f = 0.7 \text{ V}$ and $R_{dio} = 100 \text{ m}\Omega$.

Calculations: $P_{d,c} = 650 \text{ mW}$.

The reverse recovery losses are negligible because the diode current drops to zero at the end of the period time.

The forward recovery losses are negligible too because the diode current at switch-on is rather low for the applied diode.

4.6.4 Buck Capacitor

The output capacitor of the buck converter C1 is a film capacitor of the 379 MKP series of Philips. The differential resistance of the lamp is rather small and negative. To obtain a stable operating point, an

impedance in series with the lamp is needed, so an ideal voltage source is forbidden. As a result the output capacitor should be small. The film capacitor is preferred to the usual electrolytic capacitor for life-time reasons.

4.7 Control Loop

The overall control loop consists of four blocks, namely:

- The power conversion transfer (buck converter and commutator)
- The measuring circuit transfer function (R_u , R_i and C11)
- The OTA transfer (IC1 and C9)
- The modulator transfer (IC1, R_s , R4 and R5)

The power conversion transfer function is rather complex, so a quantitative description is not given at this moment. As a consequence, it does not make sense to describe the other transfer functions in a quantitative way either. So all control blocks will not be described in a quantitative way but in a general way.

Furthermore, instantaneous values are regarded to be averaged values in one HF period and the corner frequency in the feedback loop will be much lower than the operating frequency of the buck converter. So the given description is an averaged model description and is valid for rather slow variations in the circuit.

4.7.1 Power Conversion

The power conversion circuit behaves like a controlled current source due to the SOPS operation mode. The peak value of the buck coil current is controlled by the feedback loop and the minimum value will always be zero, so the average current will be equal to half the peak value, see figure 17 on page 29. Normally, the transfer function of these kind of systems are first order functions formed by C1, the lamp resistance and the current source through L1. The igniter coil L3 changes this system into a second order transfer function.

4.7.2 Measuring Circuit

The measuring circuit as described in section 4.2 is extended with an extra capacitor C11 at pin 1 of IC1. The capacitor is added to suppress the oscillations on C1 caused by the commutator circuit during the run-up phase, see section 5.1. C11 gives an extra pole in the feedback circuit and makes obtaining stability more difficult.

4.7.3 OTA in IC1

The OTA is used to set the static error to zero by integrating capacitance C9 (disregarding the tolerances on components). The corner frequency must be much lower than the operating frequency of the buck converter.

4.7.4 Modulator in IC1

The modulator in IC1 is a comparator which makes the inductor peak value, measured by R_S and tapped by R_4 and R_5 , equal to the control voltage of the OTA. This function has no pole.

Finally, figure 3 gives the Nyquist diagram of the open loop transfer function simulated with PSTAR. The parameters are the lamp resistance and the integration capacitance C_9 of the OTA at pin 2 of IC1.

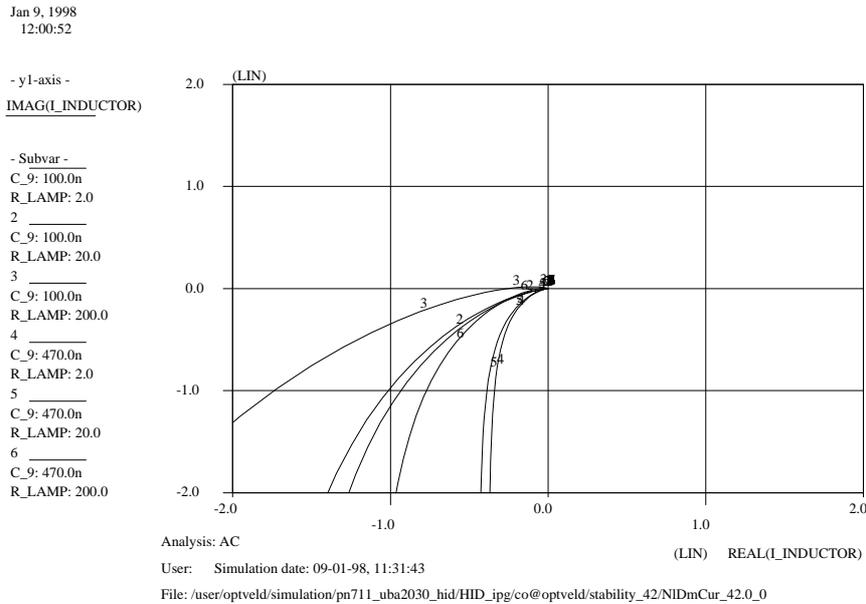


Fig.3 open loop transfer with parameters C_9 and the lamp resistance

The system is stable for all parameter combinations because the point of oscillation $(-1,0)$ is not enclosed. But the phase margin in situation 3 is rather small and will produce an overshoot in the power response.

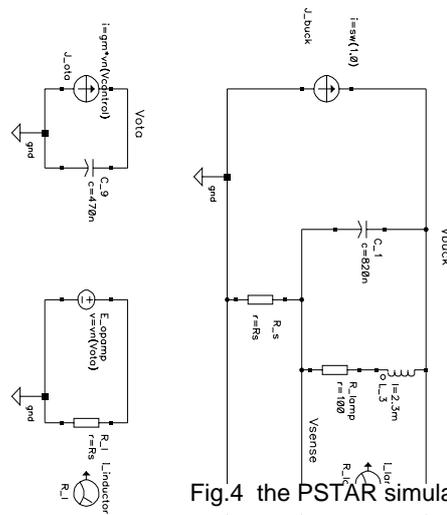


Fig.4 the PSTAR simulation circuit

5. COMMUTATOR & IGNITER.

5.1 Commutator

The commutator is a full bridge converter driven by the driver IC3 UBA2030. The diagonally transistor pair TR2 and TR5 switch simultaneously and the diagonally transistor pair TR3 and TR4 switch simultaneously either. Both transistor pairs switch alternately at a commutation frequency of ± 100 Hz, fixed by R26 (100 k Ω) and C16 (33 nF) connected to the UBA2030. The commutation frequency is asynchronous to the buck converter frequency. Locking both frequencies together is difficult and makes no difference. The dead-time is fixed at 1.7 μ s by resistor R27 (330 k Ω). The desired dead-time t_{dt} needed for proper switching is given in equation 12, with the preassumption that the maximum lamp voltage $U_{Ia,max}$ is 105 V. The nominal lamp voltage U_n is 85 V, the nominal lamp power P_n is 70 W and the dvdt limiting capacitor C_{dvdt} is 10 nF (C3 and C4).

$$\left. \begin{aligned} t_{dt} &= \frac{C_{dvdt} \cdot U_{Ia}}{I_{Ia}} \\ I_{Ia} &= \frac{P_n}{U_n} \cdot \left(2 - \frac{U_{Ia}}{U_n}\right) \end{aligned} \right\} \Rightarrow t_{dt} = \frac{C_{dvdt} \cdot U_{Ia} \cdot U_n}{P_n \cdot \left(2 - \frac{U_{Ia}}{U_n}\right)} \quad (12)$$

The dvdt limiting capacitor C_{dvdt} (C3 and C4) is determined by the maximum permissible dvdt at the pin of the IC and the maximum current through the dvdt limiting capacitor. The maximum current is about 30 A during ignition and the maximum permissible dvdt at the driver IC is 4 V/ns so that a minimal $C_{dvdt} = 7.5$ nF is desired. The applied value for C3 and C4 is 10 nF.

The voltage supply for the UBA2030 is realised by connecting the HV-pin, pin 13, directly to the buck output voltage.

The igniter coil L3 and buck output capacitor C1 will oscillate at each commutation due to the relative low value of C1, see figure 5 on page 21. Just before commutation when the system is in steady-state, the current through the lamp and L3 is equal to the averaged current delivered by the buck converter hence, the resulting current through C1 is zero and the voltage across C1 is equal to the lamp voltage. Just after commutation, the commutator switches the load in reverse so that initially twice the steady-state lamp current will flow through C1 which initiates an oscillating current through C1 and corresponding voltage across C1. The oscillation is damped by the lamp resistance. Especially during the run-up phase when the lamp resistance is very low, the oscillation is very large compared to the averaged lamp voltage which would force the buck output voltage to negative values. The negative output voltage is clamped by the body diode of the fullbridge power mosfets. The oscillation during the run-up phase causes a small gap in the voltage across C1 after each commutation, see for a detailed description section 4.1. Figure 8 on page 22 gives the voltage across C1 during the run-up phase and figure 6 on page 21 during the burn phase.

5.2 Igniter

The igniter is formed by C2, D2, L2, L3 and Z4 through Z7. The igniter generates pulse voltages of about 5000 V peak with a repetition frequency equal to the commutation frequency, ± 100 Hz. C2, L2 and L3 take care of the pulsating character of the igniter. The key components are;

- D2 is a break-over diode (sidac) type K1V36W manufactured by Shindengen with a typical break-over voltage $U_{bo} = 360$ V. This device is symmetrical in both directions.
- L2 is a fixed inductor of 270 μ H which limits the current through the break-over diode and saturates at 2 A. After saturation the inductor value is reduced to 20 μ H.
- The igniter coil L3 is an EF32 core with 3C85 as core material, primary 7 turns 0.2 mm and secondary 210 turns 0.45 mm solid copper wire. The secondary winding, connected in series with the lamp, has an inductance of 6 mH. The maximum applied voltage between the windings mutual and between the windings and core is 5600 V.
- Z4 through Z7 are zener diodes of the philips type BZD23C47C who limit the voltage across the igniter coil to prevent corona.

6. WAVEFORMS.

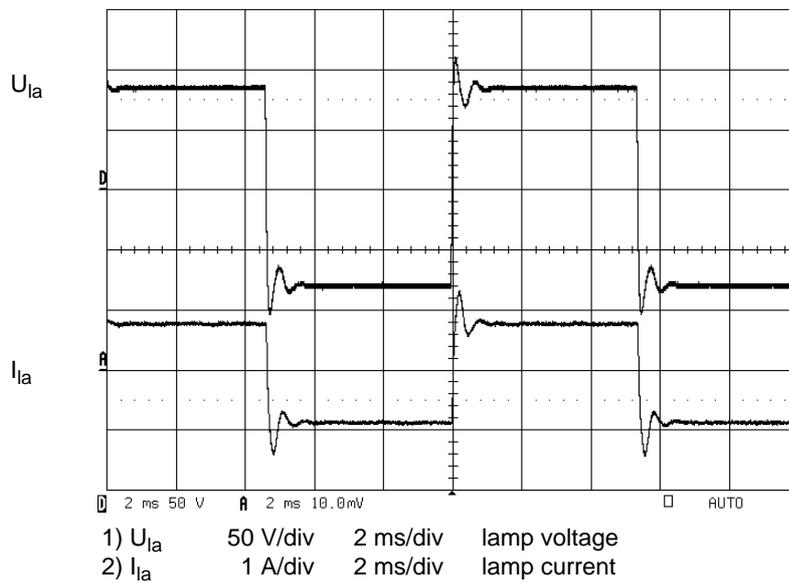


Fig.5 lamp voltage and lamp current, burn phase

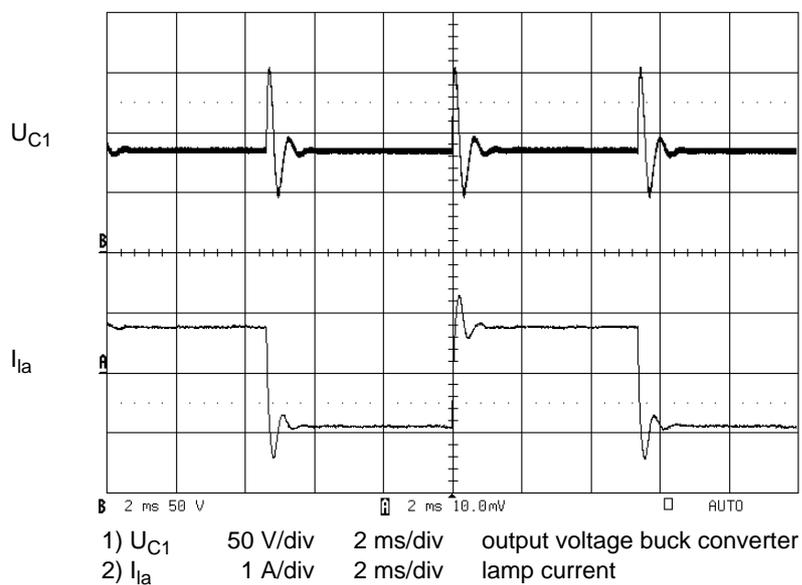


Fig.6 buck output voltage and lamp current, burn phase

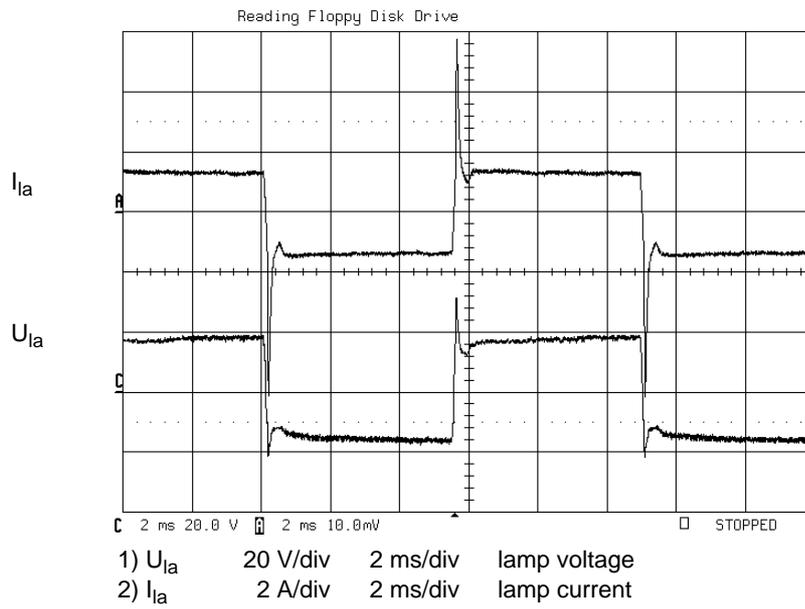


Fig.7 lamp voltage and lamp current, run-up phase

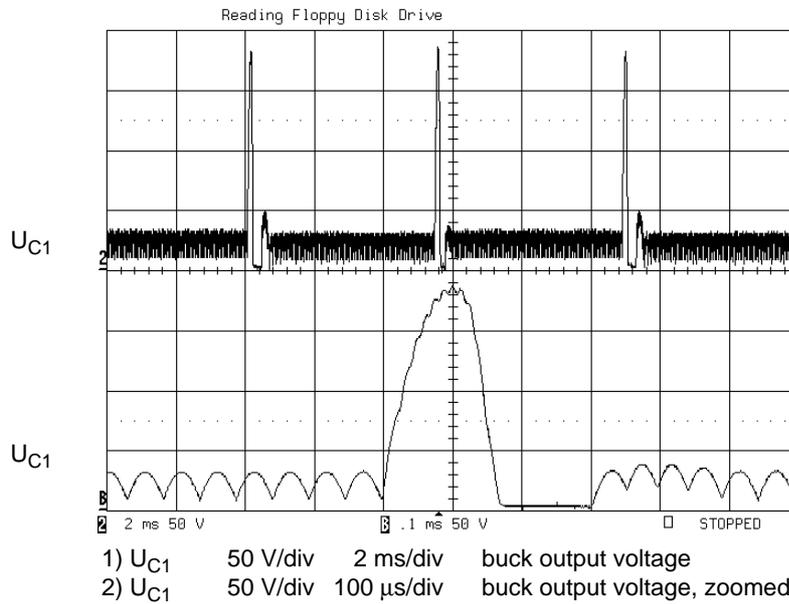


Fig.8 buck output voltage, run-up phase

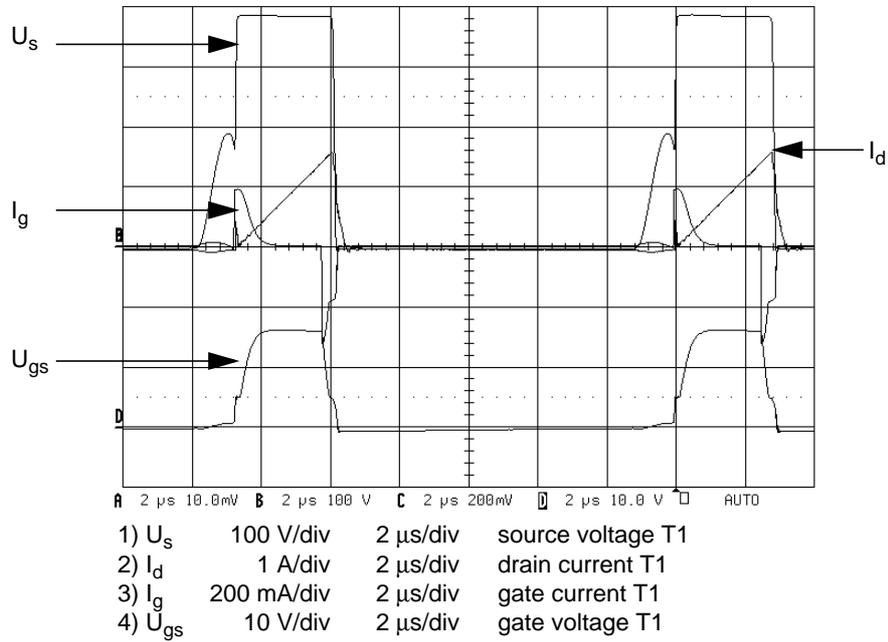


Fig.9 switch behaviour of T1

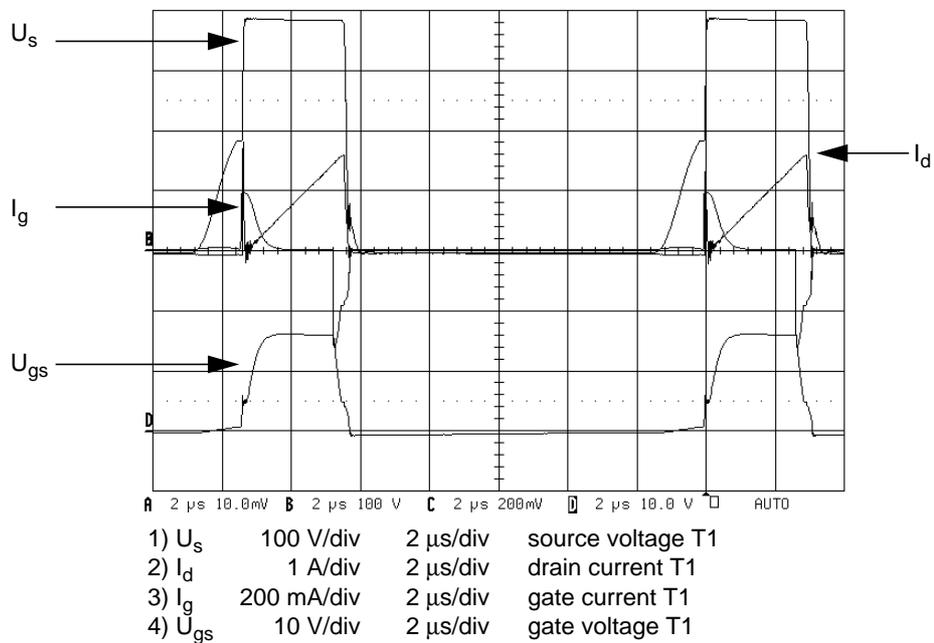


Fig.10 switch behaviour of T1, with extra $C_{ext} = 270$ pF

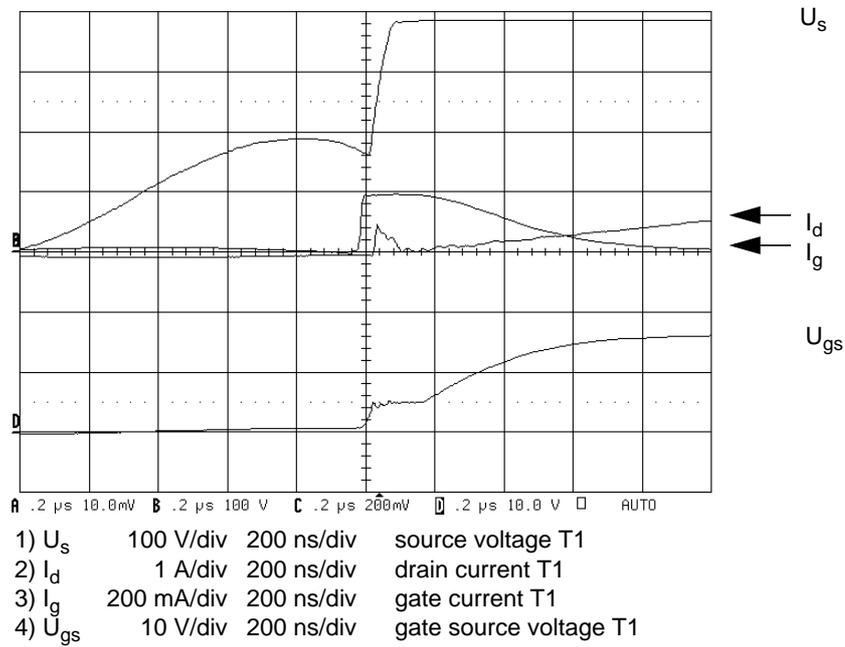


Fig.11 switch-on behaviour of T1

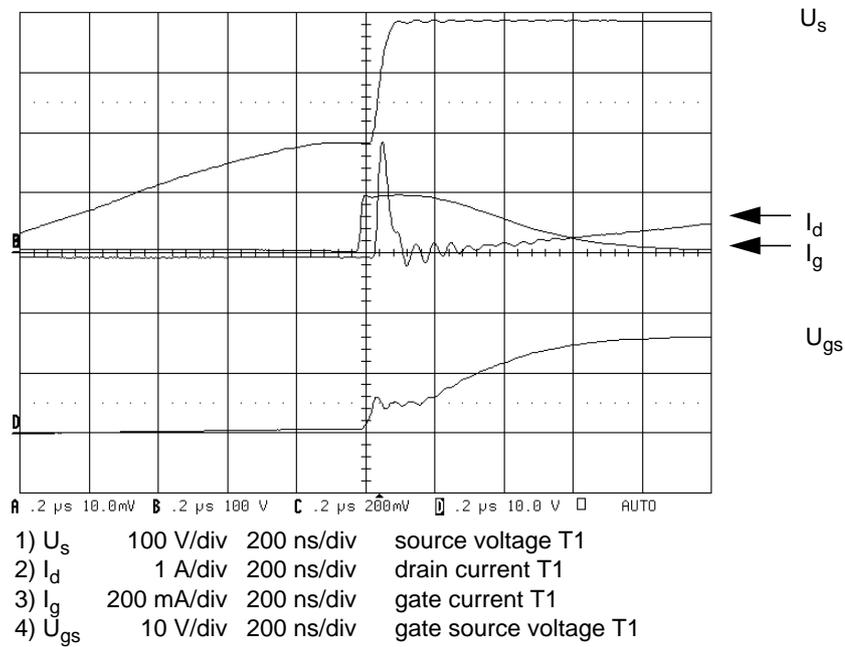


Fig.12 switch-on behaviour of T1, with extra $C_{ext} = 270$ pF

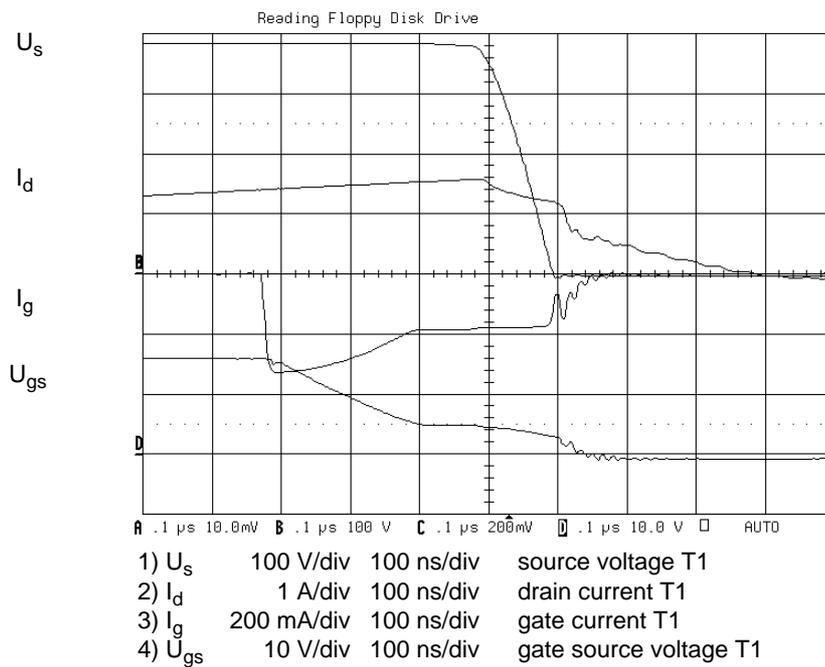


Fig.13 switch-off behaviour of T1

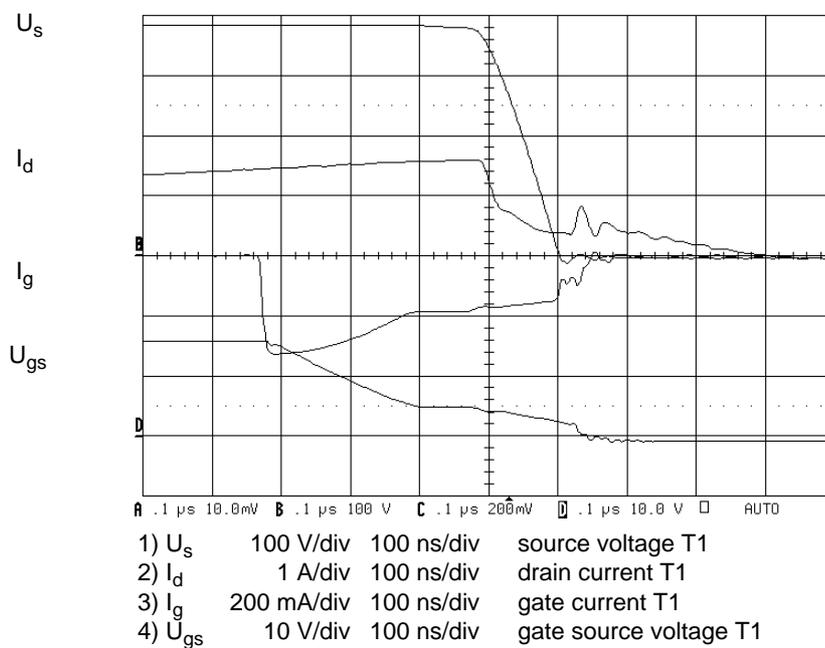


Fig.14 switch-off behaviour of T1, with extra $C_{ext} = 270$ pF

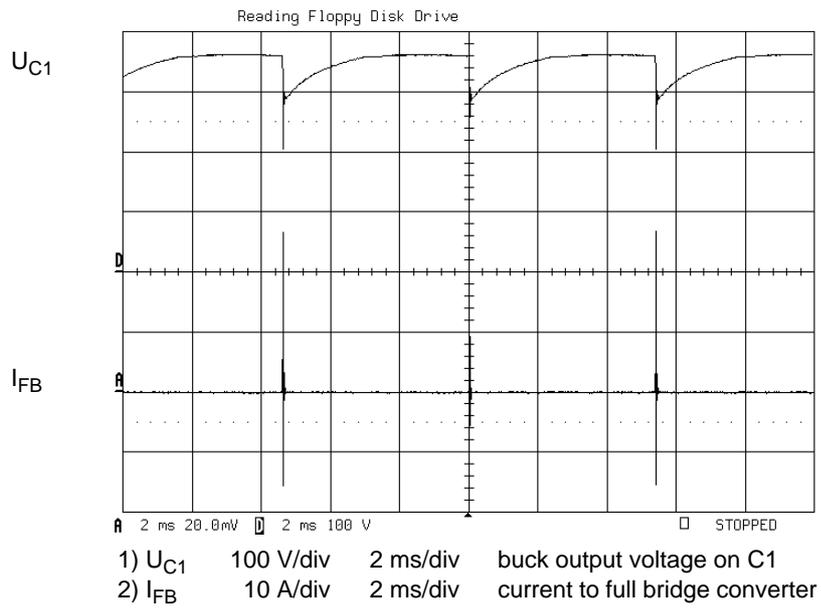


Fig.15 the voltage U_{C1} during ignition

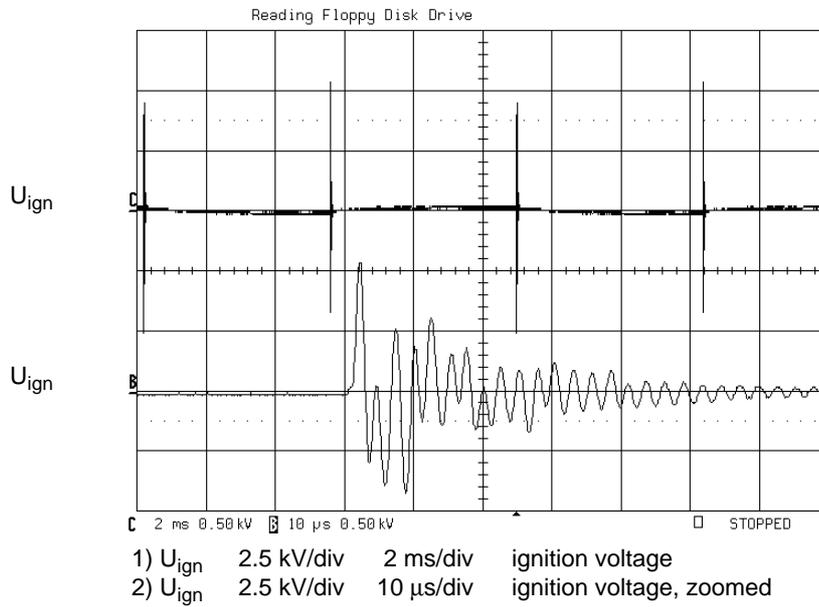


Fig.16 the secondary ignition voltage at the igniter coil L3

7. CONCLUSIONS & RECOMMENDATIONS.

Conclusions

- The ignition coil takes care for a second order transfer of the power circuit and an extra pole needed in the control circuit. So, the overall transfer is a fourth order one and makes loop stability difficult.
- An auxiliary winding on the buck coil is needed for the floating supply of the level shift IC in unloaded situations.
- The control IC1 is not completely suited for the application.

Recommendations

- The igniter circuit should be redesigned to get a smaller value of the inductance in series with the lamp.
- A lamp with ignition electrode makes the ignition coil in series with the lamp obsolete and ignition could be done by a piezoelectric transformer.

APPENDIX 1 CALCULATIONS.

In this appendix, some formulas are derived to give a mathematical description of the buck converter. The switching frequency always will be very high in relation to the commutation frequency of the full bridge converter and the applied corner frequency in the feedback circuit. The buck converter operates in a SOPS mode what means that the current through the buck coil is continues and alternates between zero and \hat{I}_L , so the average coil current $\bar{I}_L = \frac{1}{2} \cdot \hat{I}_L$ see Fig.17.

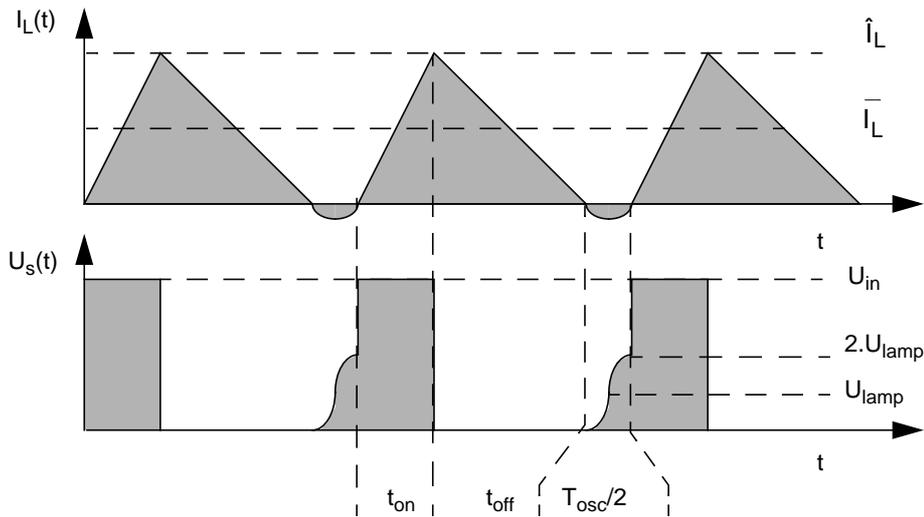


Fig.17 SOPS mode

The negative part drawn in the current in figure 17 is not according the weight scale.

A.1.1 Current Sensing

The maximum current \hat{I}_L through the buck coil (L1) and the maximum lamp current $I_{la,max}$ are given by equation 13. $U_{th,max}$ is the internal voltage limiter of the control IC.

$$\left. \begin{aligned} \hat{I}_L &= k \cdot \frac{U_{th,max}}{R_s} \\ k &= 1 + \frac{R_5}{R_4} \end{aligned} \right\} \Rightarrow \dots I_{la,max} = \bar{I}_L = \frac{\hat{I}_L}{2} = \frac{k}{2} \cdot \frac{U_{th,max}}{R_s} \quad (13)$$

A.1.2 Power Control Management

The feedback signal is obtained by adding the lamp voltage U_{la} and lamp current I_{la} . The error amplifier of control IC1 is an OTA. The output is connected to a relative large capacitor C9 so that the corner frequency f_c of the control circuit is very low compared to the switching frequency.

The inputs of the error amplifier must be equal, assuming that the control system is in steady state condition. So the voltage at the negative input of the error amplifier must be equal to the internal reference voltage $U_{ref} = 2.5 \text{ V}$ at the positive input of the amplifier. Now, a simple expression can be

found between the lamp power and lamp voltage which is given in equation 14 through 16, where R_u is the voltage feedback resistor, R_i the current feedback resistor and R_s the measuring resistor for the current feedback.

$$\left. \begin{aligned} \sum_k I_k = 0 \Rightarrow \dots \frac{U_{la} - U_{ref}}{R_u} + \frac{\bar{I}_{la} \cdot R_s - U_{ref}}{R_i} = 0 \\ \bar{I}_{la} = \frac{P_{la}}{U_{la}} \end{aligned} \right\} \Rightarrow \dots \frac{U_{la} - U_{ref}}{R_u} + \frac{P_{la}}{U_{la}} \cdot \frac{R_s - U_{ref}}{R_i} = 0 \quad (14)$$

$$\left. \begin{aligned} P_{la} = \frac{U_{la}}{R_s} \cdot \frac{R_i}{R_u} \cdot \left[\left(1 + \frac{R_u}{R_i} \right) \cdot U_{ref} - U_{la} \right] \\ \text{Define, } \dots \alpha = \frac{R_u}{R_i} \end{aligned} \right\} \Rightarrow \dots P_{la} = \frac{U_{la}}{R_s \cdot \alpha} \cdot [(1 + \alpha) \cdot U_{ref} - U_{la}] \quad (15)$$

The roots of the parabola are $U_{la} = 0 \cap U_{la} = U_{ref} \cdot (1 + \alpha)$ and the top is given by $U_{la} = \frac{U_{ref}}{2} \cdot (1 + \alpha)$.

The best power control is achieved when both the nominal lamp voltage U_n and nominal lamp power P_n are set to the top of the parabola, hence $U_n = \frac{1}{2} \cdot U_{ref} \cdot (1 + \alpha_n) \Rightarrow \alpha_n = 2 \cdot U_n / U_{ref} - 1$. Substitution of U_n into equation

15 gives the expression $P_n = \frac{U_n}{\alpha_n \cdot R_s} \cdot (2U_n - U_n) = \frac{U_n^2}{\alpha_n \cdot R_s}$. Resuming, we get some simple expressions to determine α_n and R_s which are given in equation 16.

$$\alpha_n = \frac{R_u}{R_i} = 2 \cdot \frac{U_n}{U_{ref}} - 1 \quad R_s = \frac{U_n^2}{P_n \cdot \alpha_n} \quad P_{la} = P_n \cdot \frac{U_{la}}{U_n} \cdot \left(2 - \frac{U_{la}}{U_n} \right) \quad (16)$$

The lamp current I_{la} as function of lamp voltage U_{la} is linear:

$$I_{la} = \frac{P_n}{U_n} \cdot \left(2 - \frac{U_{la}}{U_n} \right) \quad (17)$$

Observing that the maximum lamp current, limited to $I_{la,max}$ (equation 13) will restrict the interval where the linear relation $I_{la}(U_{la})$ is valid, see Fig.18. The corresponding lamp voltage at breakpoint $I_{la,max}$ is called $U_{la,min}$, see equation 18.

$$\left. \begin{aligned} I_{la} = \frac{P_n}{U_n} \cdot \left(2 - \frac{U_{la}}{U_n} \right) \\ I_{la} = I_{la,max} \\ U_{la} = U_{la,min} \end{aligned} \right\} \Rightarrow U_{la,min} = U_n \cdot \left(2 - \frac{I_{la,max} \cdot U_n}{P_n} \right) \quad (18)$$

Now, the parabolical relation $P_{la}(U_{la})$ is not valid anymore for lamp voltages lower than $U_{la,min}$. Substitution of $U_{la,min}$ in equation 16 gives the corresponding lamp power P_{br} at the breakpoint:

$$\begin{cases} I_{Ia} = \frac{P_n}{U_n} \cdot \left(2 - \frac{U_{Ia}}{U_n}\right) & U_{Ia} \geq U_{Ia, \min} \\ I_{Ia} = I_{Ia, \max} & U_{Ia} \leq U_{Ia, \min} \end{cases} \quad (20)$$

$$\begin{cases} P_{Ia} = P_n \cdot \frac{U_{Ia}}{U_n} \cdot \left(2 - \frac{U_{Ia}}{U_n}\right) & U_{Ia} \geq U_{Ia, \min} \\ P_{Ia} = U_{Ia} \cdot I_{Ia, \max} & U_{Ia} \leq U_{Ia, \min} \end{cases}$$

A.1.3 Watch-dog Timer

The watch-dog time delay t_{wd} is an ordinary first order network with time constant $\tau_{wd} = R_{wd} \cdot C_{wd}$. The unloaded capacitor is charged by a resistor to the system reference voltage $U_{ref(s)}$ and generates a high to low transition to the control IC which turns the buck mosfet on. Equation 21 gives the watch-dog delay time t_{wd} .

$$U_{ref(s)} = E \cdot \left(1 - e^{-\frac{t_{wd}}{\tau_{wd}}}\right) \Rightarrow \dots t_{wd} = \tau_{wd} \cdot \ln\left(\frac{E}{E - U_{ref(s)}}\right) \quad (21)$$

A.1.4 General Characteristics Buck Converter

The operating frequency f_{op} of the buck converter is determined by the value of the buck coil L, the input voltage U_{in} , the output voltage U_{Ia} , the output power P_{Ia} and the total parasitic capacitance at the source of the mosfet C_{par} . The latter determines the freewheel oscillation period time T_{osc} of the buck coil L with the parasitic capacitance C_{par} when the current through the buck diode drops to zero. The derivation of the operating period time T_{op} is given in equation 22.

$$\left. \begin{aligned} t_{off} &= \frac{L \cdot \hat{I}_L}{U_{Ia}} \\ t_{on} &= \frac{L \cdot \hat{I}_L}{U_{in} - U_{Ia}} \\ \hat{I}_L &= \frac{2 \cdot P_{Ia}}{U_{Ia}} \end{aligned} \right\} \Rightarrow T_{op} = \frac{L \cdot \hat{I}_L}{U_{Ia}} + \frac{L \cdot \hat{I}_L}{U_{in} - U_{Ia}} + \frac{T_{osc}}{2} = L \cdot \frac{2 \cdot P_{Ia}}{U_{Ia}} \cdot \frac{U_{in}}{U_{Ia} \cdot (U_{in} - U_{Ia})} + \frac{T_{osc}}{2} \quad (22)$$

The operating frequency f_{op} is given in equation 23 where T_{osc} is assumed to be small in relation to $(t_{on} + t_{off})$. So, during the ignition phase equation 23 is not valid (U_{Ia} is large and P_{Ia} is small).

$$f_{op} = \frac{U_{Ia}^2}{2 \cdot L \cdot P_{Ia}} \cdot \left(1 - \frac{U_{Ia}}{U_{in}}\right) \quad (23)$$

Combining equation 23 and 16 eliminates the lamp power P_{Ia} so that f_{op} is merely a function of U_{Ia} . Equation 24 gives f_{op} as function of U_{Ia} and is only valid when the lamp current is not limited.

$$\left. \begin{aligned} f_{op} &= \frac{U_{Ia}^2}{2 \cdot L \cdot P_{Ia}} \cdot \left(1 - \frac{U_{Ia}}{U_{in}}\right) \\ P_{Ia} &= P_n \cdot \frac{U_{Ia}}{U_n} \cdot \left(2 - \frac{U_{Ia}}{U_n}\right) \end{aligned} \right\} \Rightarrow f_{op} = \frac{U_{Ia} \cdot U_n}{2 \cdot L \cdot P_n} \cdot \frac{1 - \frac{U_{Ia}}{U_{in}}}{2 - \frac{U_{Ia}}{U_n}} \quad (24)$$

Furthermore, we can derive the relation between the output voltage U_{Ia} and the input voltage U_{in} . Assuming that the freewheel oscillation period time T_{osc} is small compared to $(t_{on} + t_{off})$ again, an expression of $U_{Ia}(U_{in})$ is given in equation 25.

$$\left. \begin{aligned} \hat{I}_L &= \frac{t_{off} \cdot U_{Ia}}{L} \\ \hat{I}_L &= \frac{t_{on} \cdot (U_{in} - U_{Ia})}{L} \end{aligned} \right\} \Rightarrow U_{Ia} = \frac{t_{on}}{T_{op}} \cdot U_{in} = \delta \cdot U_{in} \quad (25)$$

A.1.5 Power Mosfet Losses

The power mosfet losses are divided into three parts:

- Conduction losses $P_{m,c}$ caused by the RMS current through the mosfet I_m .
- Switch-on losses $P_{m,son}$ caused by the voltage across the mosfet at turn-on.
- Switch-off losses $P_{m,soff}$ caused by the current through the mosfet at turn-off

power mosfet conduction losses

The power mosfet RMS current I_m and the mosfet conduction losses $P_{m,c}$ are:

$$\left. \begin{aligned} I_m^2 &= \frac{1}{T_{op}} \cdot \int_0^{t_{on}} \left(\frac{\hat{I}_L}{t_{on}} \cdot t\right)^2 \cdot dt = \frac{t_{on}}{T_{op}} \cdot \frac{\hat{I}_L^2}{3} = \delta \cdot \frac{\hat{I}_L^2}{3} \\ \delta &= \frac{U_{Ia}}{U_{in}} \\ \hat{I}_L &= 2 \cdot I_{Ia} = 2 \cdot \frac{P_{Ia}}{U_{Ia}} \end{aligned} \right\} \Rightarrow P_{m,c} = R_{ds(on)} \cdot I_m^2 = \frac{4}{3} \cdot R_{ds(on)} \cdot \frac{P_{Ia}^2}{U_{Ia} \cdot U_{in}} \quad (26)$$

power mosfet switch-on losses

The switch-on losses $P_{m,son}$ are given in equation 28. T_{osc} is the freewheel oscillation at the moment that the diode current is zero. The buck coil L and the total parasitic capacitance C_{par} from source terminal to ground determine the oscillation period time T_{osc} . The oscillation period time T_{osc} is

measured so that C_{par} can be calculated. The circuit is designed to switch-on at the moment that the voltage across the mosfet is minimal, i.d. $U_{ds, son} = U_{in} - 2 \cdot U_{la}$ and the voltage across the capacitor is maximal, $U_{s, son} = 2 \cdot U_{la}$. The energy dissipated in the mosfet $E_{m, son}$ is given in equation 27.

$$E_{m, son} = \int_0^{\infty} U_{ds} \cdot I_{ds} \cdot dt = \int_0^{\infty} (U_{in} - U_s) \cdot C_{par} \cdot \frac{\partial U_s}{\partial t} \cdot dt = C_{par} \cdot \int_0^{\infty} (U_{in} - U_s) \cdot dU_s \left. \begin{array}{l} \\ U_s(0) = U_{s, so} \cap U_s(\infty) = U_{in} \end{array} \right\} \Rightarrow \dots \quad (27)$$

$$E_{m, son} = \frac{1}{2} \cdot C_{par} \cdot (U_{in} - U_{s, so})^2$$

Equation 28 gives the switch-on losses in the mosfet.

$$P_{m, son} = f_{op} \cdot E_{m, son} = f_{op} \cdot \frac{1}{2} \cdot C_{par} \cdot (U_{in} - 2 \cdot U_{la})^2 \quad (28)$$

$$T_{osc} = 2 \cdot \pi \cdot \sqrt{L \cdot C_{par}} \Rightarrow C_{par} = \left(\frac{T_{osc}}{2 \cdot \pi} \right)^2 \cdot \frac{1}{L}$$

power mosfet switch-off losses

The switch-off behaviour of the mosfet is determined by:

- the peak current through the conduction channel just before switch-off I_{ch} which equals the buck inductor peak current \hat{I}_L .
- the gate sink current during turn-off I_g .
- the transconductance of the mosfet g_m .
- the capacitors C_{gs} , C_{dg} and C_{ds} of the mosfet and some external capacitors formed by the diode capacitance, the buck coil capacitance and the parasitic capacitance of the source terminal to ground.

Some quantities are non-linear like C_{ds} , C_{dg} and g_m . For simplicity, all quantities are supposed to be linear. Now, the switch-off model description becomes an approximation of the real switch-off behaviour. Furthermore, g_m is supposed to be very large which means that a large channel current variation ΔI_{ch} is obtained by a small change in gate-source voltage ΔU_{gs} . So, U_{gs} is assumed to be constant during switch-off and will be a little bit larger than the threshold voltage of the mosfet U_{th} . Figure 19 gives the gate-source voltage U_{gs} , the drain source-voltage U_{ds} and the current through the drain-source channel I_{ch} in case of previous assumptions. The current through the channel during switch-off is denoted by I_{so} . The gate-source capacitance is not drawn in figure 19 because it does not affect the switch-off losses (U_{gs} is almost constant). Equation 29 gives the relation between I_p and I_g .

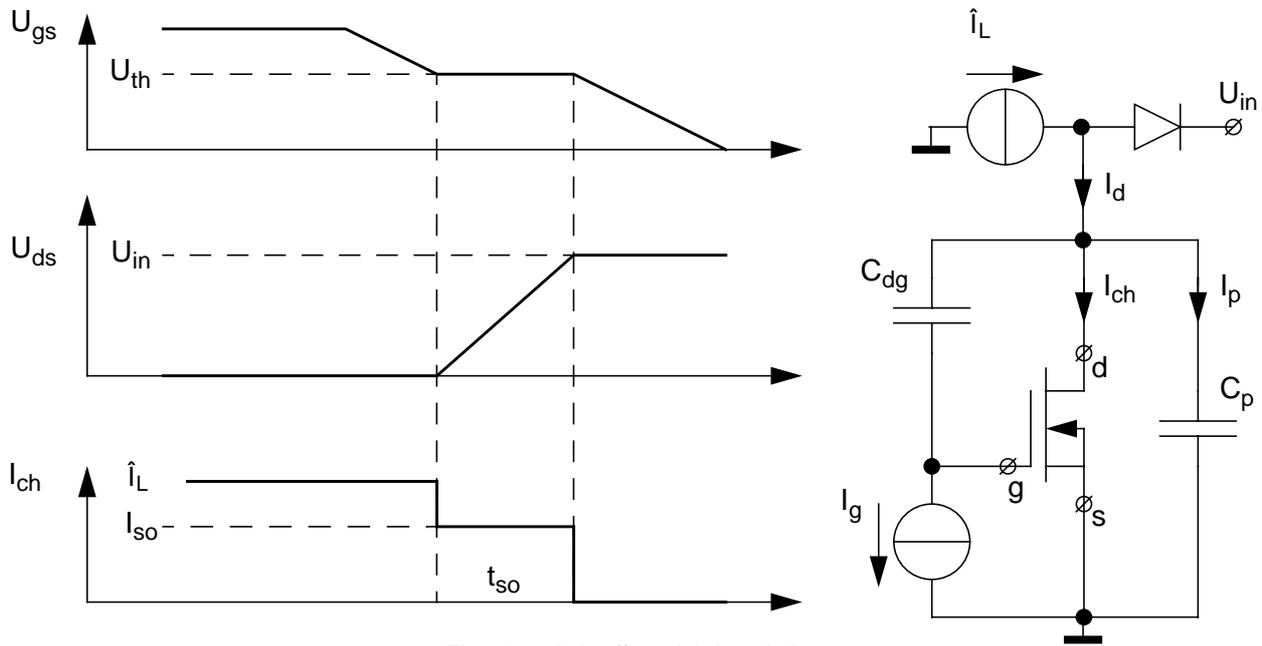


Fig.19 switch-off model description

$$\left. \begin{aligned}
 \frac{\partial U_{dg}}{\partial t} &= \frac{I_g}{C_{gd}} \\
 \frac{\partial U_{ds}}{\partial t} &= \frac{\partial(U_{dg} + U_{gs})}{\partial t} = \frac{\partial U_{dg}}{\partial t} \\
 \frac{\partial U_{ds}}{\partial t} &= \frac{I_p}{C_{ds} + C_{ext}}
 \end{aligned} \right\} \Rightarrow \frac{I_g}{C_{gd}} = \frac{I_p}{C_p} \Rightarrow I_p = \frac{C_p}{C_{gd}} \cdot I_g \quad (29)$$

The drain-source channel current during switch-off I_{so} , the drain-source voltage U_{ds} and the switch-off time t_{so} are given by equation 30, where k is used to reduce the formula length.

$$\begin{aligned}
 I_{so} &= \hat{I}_L - (I_g + I_p) = \hat{I}_L - \left(1 + \frac{C_p}{C_{gd}}\right) \cdot I_g = \hat{I}_L - k \cdot I_g \\
 \frac{\partial U_{ds}}{\partial t} &= \frac{\partial U_{dg}}{\partial t} = \frac{I_g}{C_{gd}} \Rightarrow U_{ds}(t) = \frac{I_g}{C_{gd}} \cdot t \\
 U_{ds}(t_{so}) &= \frac{I_g}{C_{gd}} \cdot t_{so} = U_{in} \Rightarrow t_{so} = \frac{U_{in} \cdot C_{gd}}{I_g}
 \end{aligned} \quad (30)$$

The switch-off losses are given by multiplication of the drain-source channel current I_{so} and the drain-source voltage U_{ds} during the switch-off time t_{so} :

$$\begin{aligned}
P_{m, \text{soff}} &= \frac{1}{T} \cdot \int_0^{t_{\text{so}}} U_{\text{ds}}(t) \cdot I_{\text{so}} \cdot dt = \frac{1}{T} \cdot \int_0^{t_{\text{so}}} \frac{I_g}{C_{\text{gd}}} \cdot t \cdot (\hat{I}_L - k \cdot I_g) \cdot dt \\
P_{m, \text{soff}} &= \frac{1}{T} \cdot (\hat{I}_L - k \cdot I_g) \cdot \frac{I_g}{C_{\text{gd}}} \cdot \frac{1}{2} \cdot t_{\text{so}}^2 \\
P_{m, \text{soff}} &= \frac{1}{T} \cdot (\hat{I}_L - k \cdot I_g) \cdot \frac{I_g}{C_{\text{gd}}} \cdot \frac{1}{2} \cdot \left(\frac{U_{\text{in}} \cdot C_{\text{gd}}}{I_g} \right)^2 = f \cdot \frac{1}{2} \cdot C_{\text{gd}} \cdot U_{\text{in}}^2 \cdot \left(\frac{\hat{I}_L}{I_g} - k \right)
\end{aligned} \tag{31}$$

Be careful: equation 31 is only valid when $\hat{I}_L > k \cdot I_g$ is fulfilled, otherwise the switch-off losses are zero.

A.1.6 Diode Losses

The diode conduction losses $P_{\text{d,c}}$ will be a first order approximation, see equation 32, with U_f as diode forward voltage and R_{dio} as diode resistance. Both values are measured on a curve tracer. The conduction losses are expressed in the average diode current and RMS diode current.

$$\begin{aligned}
P_{\text{d,c}} &= \frac{1}{T} \cdot \int_0^T u(t) \cdot i(t) \cdot dt = \frac{1}{T} \cdot \int_0^T [U_f + R_{\text{dio}} \cdot i(t)] \cdot i(t) \cdot dt = \frac{1}{T} \cdot \int_0^T U_f \cdot i(t) \cdot dt + \frac{1}{T} \cdot \int_0^T R_{\text{dio}} \cdot i(t)^2 \cdot dt \\
P_{\text{d,c}} &= U_f \cdot \bar{I}_d + R_{\text{dio}} \cdot I_d^2
\end{aligned} \tag{32}$$

The RMS current, see equation 33, is similar to the derivation of the mosfet RMS current shown in equation 26.

$$\left. \begin{aligned}
I_d^2 &= \frac{1}{T_{\text{op}}} \cdot \int_0^{t_{\text{off}}} \left(\frac{\hat{I}}{t_{\text{off}}} \cdot t \right)^2 \cdot dt = \frac{t_{\text{off}}}{T_{\text{op}}} \cdot \frac{\hat{I}^2}{3} = (1 - \delta) \cdot \frac{\hat{I}^2}{3} \\
\delta &= \frac{U_{\text{Ia}}}{U_{\text{in}}} \\
\hat{I} &= 2 \cdot I_{\text{Ia}} = 2 \cdot \frac{P_{\text{Ia}}}{U_{\text{Ia}}}
\end{aligned} \right\} \Rightarrow I_d^2 = \frac{4}{3} \cdot \left(\frac{U_{\text{in}}}{U_{\text{Ia}}} - 1 \right) \cdot \frac{P_{\text{Ia}}^2}{U_{\text{Ia}} \cdot U_{\text{in}}} \tag{33}$$

The average diode current is equal to the lamp current, so the approximated conduction losses $P_{\text{d,c}}$ are:

$$P_{\text{d,c}} = U_f \cdot \frac{P_{\text{Ia}}}{U_{\text{Ia}}} + R_{\text{dio}} \cdot \frac{4}{3} \cdot \left(\frac{U_{\text{in}}}{U_{\text{Ia}}} - 1 \right) \cdot \frac{P_{\text{Ia}}^2}{U_{\text{Ia}} \cdot U_{\text{in}}} \tag{34}$$

APPENDIX 2 COMPLETE CIRCUIT DIAGRAM.

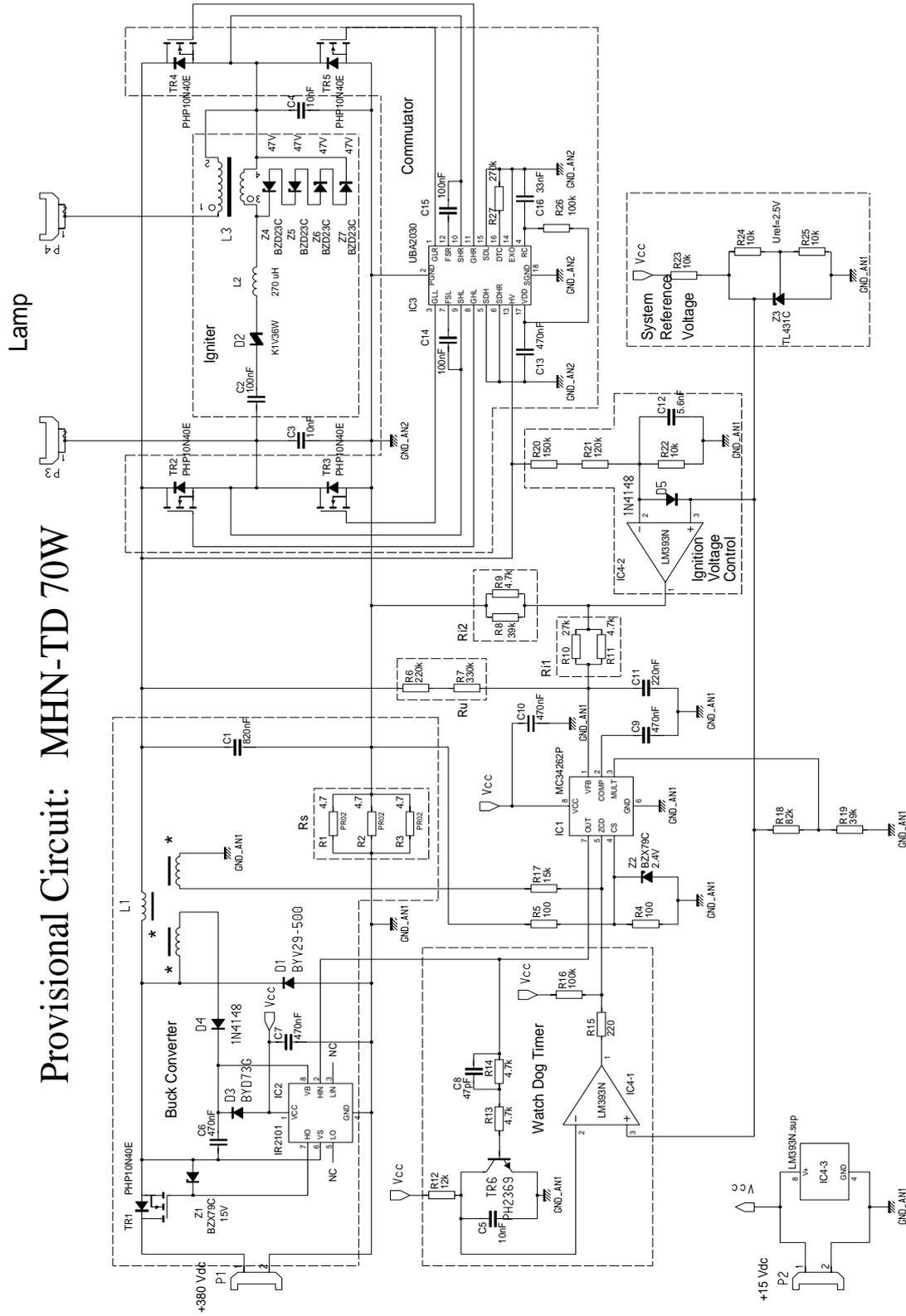


Fig.20 complete circuit diagram

