A low cost electronic ballast for a 50V MHS gas-discharge lamp used for car headlights

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Award date:
1994
A LOW COST ELECTRONIC BALLAST
FOR A 50V MHS GAS-DISCHARGE
LAMP USED FOR CAR HEADLIGHTS

EMV 94-08     Jeroen Kleinpenning

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Eindhoven   : 10 juni 1994
REPORT EMV 94-08

TITLE: A low cost electronic ballast for a 50V MHS gas-discharge lamp used for car headlights

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DATE: 1994-06-10

PROJECT: Automotive, MPL Quick Start Ballast, generation 3

COMPANY: Philips Lighting

DEPARTMENT: Development Lighting Electronics Eindhoven

LANGUAGE: English

KEY WORDS: Electronic ballast, metal halide lamp, MHS lamp, car headlights, DC-DC power conversion, boost converter, tapped-inductor, commutator, ignitor

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SUMMARY

A miniaturized gas-discharge lamp has been designed for a gas-discharge light system for car headlights. The photometric performance of such a high-pressure gas-discharge lamp is superior in comparison with a filament halogen lamp. However, the gas-discharge lamp requires an expensive electronic ballast and ignitor.

The goal was to design a new ballast and ignitor with a low price and a small volume. The designed low cost ballast is based on a metal halide sodium scandium lamp with 50V lamp voltage and 35W nominal lamp power. With an increased run-up power (100W) the 25% level of the nominal luminous flux is reached within one second after ignition of the lamp.

The ballast is supplied by the car battery and contains a power converter, a high-voltage circuit and a commutator. Power conversion (up to 100W) is performed by a boost converter operating at 500kHz with 60V maximum output voltage and an efficiency of 85%. Ignitor voltage (800V) and the open circuit voltage for take-over and re-ignition (200V) are generated by a tapped-inductor boost converter. The lamp current is supplied via the commutator and has a low frequency (250Hz) square wave form. The commutator is a full bridge with integrated circuits for gate driving. The ignitor with a 800V spark gap as switching element generates a 25kV ignition pulse. The ignitor contains a custom-made transformer.

The component price of the power part of the ballast is DFL 12.81, the component price of the ignitor is DFL 5.57. Both prices are without costs for mounting and housing. The ballast is designed with two small magnetic components: an U11 inductor and an E13 transformer. The efficiency of the ballast is 82% for 12V battery voltage and 35W lamp power.
PREFACE

I was very pleased to have the opportunity to do my graduation work at Philips Lighting, the only company in the world able to manufacture a miniaturized gas-discharge lamp for an automotive application. I like to thank Jaap Rozenboom and Theo Stommen for offering me this graduation task, Harry Linssen for the excursion to the lamp factory in Aachen, Nico de Jong for his advice on magnetics, Theo Hendriks and John Maes for their advices on electronics and Henk van Esveld and Leon Konings for their explanation of gas-discharge lamps.

Jeroen Kleinpennings
Eindhoven, June 1994
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1 INTRODUCTION

A miniaturized gas-discharge lamp has been designed for a gas-discharge light system for car headlights. Such a system contains the following parts:
- small gas-discharge lamp,
- carefully designed reflector,
- electronic ballast and ignitor.

The advantages of a gas-discharge lamp in comparison with a filament halogen lamp are:
- higher luminous efficacy (factor 4),
- longer lifetime (factor 5),
- smaller dimensions: almost a point source, allows a compact and aerodynamic headlight design.

The gas-discharge light system has a highly accurate light distribution. So an exact definition of the light/dark boundary and a good anti-glare effect are obtained.

The disadvantage is the price, the system is expensive. In particular the ballast is very expensive.

1.1 HISTORY AND PRESENT MPL SYSTEM

Several automotive gas-discharge light systems have been developed in the past. All these ballasts supplied a 35W gas-discharge lamp and were capable of hot-restrike and instant light due to an increased run-up power.

1987, An electronic ballast for a white-son lamp:
The lamp is a high-pressure sodium lamp [2] with 32V steady-state voltage. The converter inside the ballast is a boost converter with current mode control, the open circuit voltage is 180V. Ignition occurs through a 4kV pulse, generated by a pulse ignitor with a sidac as switching element.
1989, Totally Combined System [10]:
First and successful system approach for bringing together ballast, ignitor and lamp with reflector into one single unit. The technical features fast run-up, reliable ignition and small overall dimensions had been given top priority. The lamp is a metal halide sodium scandium lamp with 45V steady-state lamp voltage. The ballast contains a boost converter with a maximum output voltage of 150V. Ignition is performed by an ignitor containing two cascaded transformers. The first transformer generates 1200V with a sidac as switching element. The second transformer with a gate turn-off thyristor generates the 20kV ignition pulse.

1992, MPL Quick Start Ballast [8,11,12]:
This is the present micro power light system with a 85V MHS lamp. The ballast contains a PWM-controlled flyback converter able to supply a voltage in a range from 20V up to the 550V open circuit voltage. The 1600V for supplying the ignitor is generated with an auxiliary winding on the flyback transformer. The ignitor with a spark gap as switching element generates a 25kV ignition pulse. The ballast contains a DC coil for re-ignition.

1.2 COMPETITOR BALLASTS

Up till now Philips does have the splendid position to be the only manufacturer in the world able to make a 35W MHS gas-discharge lamp for an automotive application. So fortunately only competitor ballasts with Philips lamps exist instead of complete competitor systems.

Bosch: The ballast manufactured by Bosch contains a microprocessor for control and a patented converter: a tapped input inductor Čuk converter [27]. Ignition occurs through a pulse ignitor with two spark gaps. The ignitor is supplied with the 350V open circuit voltage of the commutator inside the ballast. This ballast is built in the BMW 700 series.
Hughes (General Motors): The ballast produced by Hughes has a sepic converter [28] with an auxiliary 1kV winding for the ignitor supply. The converter with current mode control has coupled magnetics to reduce input current ripple [27]. The ballast contains a re-ignition coil. Due to GM politics this ballast is used in the Cadillac.

Hella: Hella uses a flyback converter. This well-known converter has an open circuit voltage of 350V and an auxiliary winding for the ignitor supply. Ignition occurs through a pulse ignitor with a 800V spark gap. Take-over is realized with a high current obtained from a controlled capacitor discharge. Remarkable is the low primary inductance of the flyback transformer.

To achieve a reliable ignition with a reduced open circuit voltage Bosch and Hella delay the first current commutations. Hella supplies the lamp 70ms with a direct current after ignition. Bosch supplies the lamp 15ms with a direct current in one direction and after the first current commutation 15ms in the opposite direction, to keep the lamp current symmetrical.

All ballasts are expensive because they contain:
- a power converter with a transformer,
- semiconductors with a high maximum voltage rating,
- a re-ignition coil (Philips and Hughes).

1.3 REQUIREMENTS FOR A NEW BALLAST

The requirements for a new ballast are a low price and a small volume for achieving a significant increase of market penetration. The new ballast has to be inexpensive!

It is only possible to achieve a low ballast price if both the lamp and the ballast can be changed (the system approach). This means a ballast powered from the car battery and a gas-discharge lamp modified to achieve a low cost ballast. In the next chapter
the MHS gas-discharge lamp is discussed and a proposal for a new lamp with lower lamp voltage (50V) is given. This lower lamp voltage creates the possibility for a ballast with a boost topology [22]. A boost converter is an inexpensive converter without a transformer (see the introduction of chapter 4).

The boost converter has a limited voltage gain and is therefore not suitable for the present 85V MHS lamp with 12V battery voltage. Power conversion at this high voltage gain can not be achieved with an inexpensive boost converter. This is explained at the end of chapter 4 in section 4.10.

In chapter 3 the power part of the ballast is presented and in the following chapters each part of the ballast is discussed. The control part of the ballast was not part of my assignment and is therefore not discussed in this report. The design of the ignitor is described in chapter 7. The performance of the low cost ballast and ignitor with low voltage lamps is described in chapter 8. Finally price calculations of the presented ballast and ignitor are given.
2 MHS GAS-DISCHARGE LAMP

The gas-discharge lamp transforms the electric energy into heat and electromagnetic radiation. A considerable part of the radiation is at visible wavelengths. The gas-discharge lamp used for car headlights is a high-pressure metal halide sodium scandium lamp with a nominal electric input power of 35W.

2.1 PRINCIPLES OF GAS-DISCHARGE

The quartz glass discharge tube with sealed-in electrodes is filled with xenon, metal halides and mercury. The electrons emitted by the electrodes are accelerated by the electric field. These free electrons can have elastic collisions or inelastic collisions with the gas atoms and molecules (Meyer and Nienhuis [1]). An elastic collision between a free electron and a gas atom leads to heat generation. A small part of the low kinetic energy of the electron will be transferred to the gas atom. The high number of collisions in a gas-discharge leads to a rise in gas temperature.
An inelastic collision between an electron with high kinetic energy and a gas atom leads to excitation of the gas atoms. The excitation energy obtained from the electron will leave the atom as electromagnetic radiation.

![Figure 2.3: Inelastic collision, excitation and radiation](image)

An inelastic collision between an electron with a still higher kinetic energy and a gas atom leads to ionization of gas atoms. Ionization increases the number of free electrons and ions. Ionization of atoms or molecules is necessary for making an electrical current in the discharge possible.

![Figure 2.4: Inelastic collision, ionization](image)

The increasing number of free electrons due to ionization leads to an unlimited electrical current through the discharge tube. To prevent this, a current limitation is required. A series impedance or an electronic ballast limits and controls the lamp current.

It is possible to obtain a high output of light from electric power (high luminous efficacy) with a small gas-discharge lamp (high luminance) by means of a high-pressure discharge.
The luminous flux of a light source is the integration of the product of the visible electromagnetic radiation and the sensitivity of our eyes over the emitted spectrum. Unit: lumen (lm). The luminous efficacy of a light source is the quotient of the luminous flux emitted and the electric power consumed. Unit: lumen per watt (lm/W).

2.2 METAL HALIDE LAMPS\(^{[1,3]}\)

A method to improve luminous efficacy and colour of a high-pressure mercury lamp is to add metal halides to the discharge. Metal halides are compounds of metals and halogens. The halogen iodine is most commonly used in metal halide lamps. The metal halides have a higher vapour pressure at a given temperature than the element itself, the compounds are more volatile than their metals. The metal halide compounds start melting and evaporating at a certain temperature of the wall of the discharge tube. The vapour is carried into the hot region of the arc by diffusion and convection, where the molecules dissociate into metal\(^+\) and halide ions. Radiation is emitted by the excited metal ions or atoms in the discharge area. The ions or atoms combine to a less-aggressive metal halide compound again when they diffuse away from the hot discharge area into the lower temperature outside areas near the wall of the discharge tube.

The excitation levels of the metal halides are much lower than those of mercury. Therefore the mercury doesn't serve as a generator of light. The mercury has become an element for the regulation of the arc voltage and the heat: it is a buffer gas. To obtain a lamp with a high luminous efficacy and a good white-light approach, metal halides of sodium and scandium are used. The sodium scandium lamp has a multi-line spectrum and a colour temperature of 4000K.
Sodium-iodide is added to improve the colour and serves as a so-called diffuser. Sodium has a low ionization potential and causes more free electrons in the low temperature region of the arc. These free electrons increase the current density in this region and the power dissipated.

The nominal lamp voltage of a MHS lamp is a result of the mercury dose and the electrode distance. The mercury is dosed in an accurate determined quantity and is fully evaporated during lamp operation.

2.3 MHS LAMP FOR CAR HEADLIGHTS

For a 35W metal halide lamp without any feature for a fast run-up, the run-up time is about one minute (run-up time is defined as the time between ignition of the lamp and the moment that 25% of the nominal luminous flux is reached). For an automotive application a fast run-up is required. This requirement can be achieved by an increased electric lamp power after ignition of the lamp and by applying a high-pressure xenon filling. The use of xenon leads to an increase of the run-up voltage, for instance from 12V to 25V and the presence of a xenon discharge with radiation partly in the visible spectrum. These properties are both very advantageous for a fast run-up. With an increased run-up power supplied to the lamp and a high-pressure xenon filling the 25% level of the nominal luminous flux can be reached within one second. This 25%/1s requirement is the Vedilis standard for the run-up (Vedilis = Vehicle Discharge Light System).

Due to this high-pressure xenon filling the lamp requires a high ignition voltage and is therefore difficult to ignite.

The energy balance of a 35W MHS D2S lamp is given in figure 2.5, 26% of the electrical energy is converted into visible radiation.
2.4 PROPOSAL FOR A 50V LAMP

The data of the present D1/D2/D2S lamps with 85V nominal lamp voltage and the data of a proposal for a new lamp are given. The new lamp has a nominal lamp voltage of 50V. This reduced lamp voltage is absolutely necessary for achieving a low ballast price and a small volume of the ballast, conform to the 'Totally Combined System' knowledge [10]. A lamp with a lower lamp voltage can be achieved by reducing the mercury dose. A mercury dose of 0.1mg results in a lamp voltage of 50V. The luminous efficacy will decrease with 8%, according to experiments from Van Esveld [5]. The optical parameters change, the arc-curvature decreases and the diffusity increases with the reduced mercury dose. These parameters work in opposite direction. The optical performance of the new lamp remains therefore virtually unchanged.
Data of the 35W MHS lamps:  D1/D2/D2S lamps  new lamp (D3?)

Quartz glass discharge tube:

<table>
<thead>
<tr>
<th></th>
<th>D1/D2/D2S</th>
<th>new lamp (D3?)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Inner diameter (mm)</td>
<td>2.7</td>
<td>2.7</td>
</tr>
<tr>
<td>Outer diameter (mm)</td>
<td>6.0</td>
<td>6.0</td>
</tr>
</tbody>
</table>

Electrodes:

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<th>D1/D2/D2S</th>
<th>new lamp (D3?)</th>
</tr>
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<tbody>
<tr>
<td>Electrode distance (mm)</td>
<td>3.9</td>
<td>3.9</td>
</tr>
<tr>
<td>Electrode diameter (μm)</td>
<td>250</td>
<td>250</td>
</tr>
<tr>
<td>- tungsten 98.5%</td>
<td></td>
<td></td>
</tr>
<tr>
<td>- thorium 1.5%</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Filling:

<table>
<thead>
<tr>
<th></th>
<th>D1/D2/D2S</th>
<th>new lamp (D3?)</th>
</tr>
</thead>
<tbody>
<tr>
<td>xenon (bar)</td>
<td>7</td>
<td>7</td>
</tr>
<tr>
<td>mercury (mg)</td>
<td>0.7</td>
<td>0.1</td>
</tr>
<tr>
<td>metal halides (mg)</td>
<td>0.2</td>
<td>0.2</td>
</tr>
<tr>
<td>- sodium-iodide 79%</td>
<td></td>
<td></td>
</tr>
<tr>
<td>- scandium-iodide 16%</td>
<td></td>
<td></td>
</tr>
<tr>
<td>- thorium-iodide 5%</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Electrical data:

<table>
<thead>
<tr>
<th></th>
<th>D1/D2/D2S</th>
<th>new lamp (D3?)</th>
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<tbody>
<tr>
<td>lamp voltage (V)</td>
<td>85</td>
<td>50</td>
</tr>
<tr>
<td>ignition voltage (kV)</td>
<td>25</td>
<td>25</td>
</tr>
<tr>
<td>run-up voltage (V)</td>
<td>25</td>
<td>25</td>
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Photometric data:

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</tr>
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<tr>
<td>luminous flux (lm)</td>
<td>3000</td>
<td>&lt; 2760</td>
</tr>
<tr>
<td>luminous efficacy (lm/W)</td>
<td>85</td>
<td>&lt; 78</td>
</tr>
<tr>
<td>colour temperature (K)</td>
<td>4200</td>
<td>4200</td>
</tr>
</tbody>
</table>

Colour impressions:

- xenon: green-white
- sodium: orange
- scandium: blue-green-white
2.5 IGNITION OF THE MHS LAMP

Before the lamp is ignited the lamp behaves like an insulator due to the insulating properties of the 7 bar xenon filling. The mercury liquid and the solid metal halides are precipitated at the cold wall of the discharge tube. The lamp is ignited by supplying a high voltage, 25kV for a hot-restrike. This high voltage can be generated by a pulse ignitor. In case the following conditions are present:
- fast risetime, 25kV in 100ns
- strong inhomogeneous electrical field near the electrode tip
- high-pressure gas filling
- certain quantity of initial electrons,
the mechanism for breakdown is a streamer, and not a Townsend mechanism. A streamer is a very fast mechanism, the first avalanche directly results in a breakdown. The present initial electrons accelerate to the anode and create an ionization avalanche due to their collisions with gas atoms, Raether [7]. In this anode directed streamer a positive space charge remains because of the less mobility of the ions. This space charge enhances the electrical field in the head of the avalanche. In case the avalanche reaches an absolute number of $10^8$ free electrons [7] there will be a streamer directed to the cathode. The foton radiation of the excited atoms produces new electrons around the avalanche head. The electrons are attracted by the positive space charge and cause new foton radiation. On account of this ionizing radiation the streamer propagates very fast towards the cathode. The total breakdown occurs in about 10ns. When the streamer reaches the cathode there is a conducting channel between the electrodes.

To sustain this conducting breakdown channel a high current (in relation to the nominal current) has to be supplied to the lamp. In case of a hot-restrike the pressure inside the lamp is high and the take-over current supplied to the lamp has to be sufficient to prevent the discharge from extinguishing. The duration of the
take-over current is in the order of 10\(\mu\)s to 100\(\mu\)s. The take-over current is followed by the run-up current. The run-up current heats the electrodes, the temperature rises and gets high enough to emit the electrons thermionically (10ms to 100ms after ignition, depending on the run-up current). The light during take-over and the first phase of a cold run-up is obtained from the xenon discharge.

2.6 RUN-UP OF THE MHS LAMP

After a successful breakdown and take-over the lamp directly enters the arc phase. The MHS lamp is characterized by the absence of a perceptible ‘glow to arc’ transition during ignition. The maximum allowable run-up current, to prevent the electrode tip from melting, is 2.6A. Minimum run-up current for hot-restrike, to prevent the discharge from extinguishing, is 1.5A. These current ratings apply to an electrode diameter of 250\(\mu\)m and are determined experimentally.

If the lamp is supplied by a direct current the positive ions will concentrate at the cathode and the electrons will concentrate at the anode. Due to the fact that the mass and therefore the heat capacitance of ions is greater than that of electrons the cathode will be much cooler than the anode. To have the electrodes both at the same temperature, below the melting temperature of tungsten (3680K), the lamp has to be supplied with an alternating current. A low frequency square wave current operation avoids difference in electrode temperature and acoustic resonance of the arc. Minimum frequency of the square wave current applied to the lamp is 250Hz.

The most critical moments for the lamp to extinguish (after a successful ignition and take-over) are the first current commutations. If the lamp extinguishes the ballast offers the open circuit voltage (200V). If the combination of the open circuit voltage and cathode temperature is high enough for emitting electrons from the cathode the lamp breaks down again because of a
Townsend mechanism (or generation mechanism) [7]. This is called re-ignition. The thermic time constants of the electrodes are in the order of milliseconds (minimum frequency of the lamp current is 250Hz), so the open circuit voltage has to be supplied within 1ms.

In case the electrodes are to cold the lamp extinguishes definitely and there has to follow a new ignition pulse to start the process ignition, take-over and run-up again.

The run-up power (product of run-up current and lamp voltage) heats the lamp. The mercury and the metal halides start to evaporate. A perceptible change of lamp colour occurs, from a xenon discharge to a metal halide discharge. At a cold run-up the orange sodium discharge is the first metal halide discharge perceptible. If all the mercury and metal halides are evaporated the lamp voltage stabilises. This steady-state condition is reached in about 10s after ignition of a cold lamp (depending on the run-up power). A MHS D2 lamp with 35W lamp power in steady-state operation is in a thermal equilibrium and has an gas pressure inside of 80 bar. The hot region of the arc has a temperature of 5000K, the electrodes have a temperature of 3000K and the upper side of the discharge tube has a temperature of 1300K (the melting temperature of quartz is 1520K).

To minimize the migration of sodium through the quartz wall of the discharge tube the lamp has to be operated with a negative voltage in respect to its ambient [5,6]. This is necessary for a long lamp life. The positive voltage of the ballast has to be connected via a high resistance to the reflector.
3 BALLAST IGNITOR LAMP SYSTEM

The ballast controls and limits the lamp power and supplies the ignitor. The ballast contains the following circuits:
- power converter: a boost converter
- high-voltage circuit: a tapped-inductor boost converter
- commutator: a full bridge

The ballast, ignitor and lamp configuration is depicted in the figure below.

![Ballast ignitor lamp configuration](image)

**Figure 3.1: Ballast ignitor lamp configuration**

3.1 SYSTEM ANALYSIS

The boost converter turns the battery voltage into a higher voltage. The boost converter supplies continuously the lamp power within a maximum lamp current of 2.6A and a maximum output voltage of 60V. Because of this voltage limit the components can be inexpensive and small in size and the power conversion occurs with high efficiency.

The high-voltage circuit generates the 800V for supplying the ignitor and the 200V open circuit voltage for the take-over and re-ignition. This HV-circuit is a tapped-inductor boost converter and is supplied by the boost converter. The HV-circuit is only active during ignition of the lamp and in case of a re-ignition.
This high-voltage circuit can therefore be built with inexpensive components since efficiency and heat generation are less important.

The commutator transforms the DC voltage, with a maximum of 200V, into a low frequency square wave voltage. The commutator supplies the current to the lamp through the series ignitor.

The ignitor is powered by the HV-circuit and generates the 25kV ignition pulse. The ignitor contains a 800V spark gap as switching element. Both the HV-circuit and the ignitor are not active during steady-state operation.

This new ballast approach with a main high efficiency converter supplying the lamp power (up to 100W) with a limited output voltage of 60V and a second inexpensive converter (5W) generating the 800V ignitor voltage and the 200V open circuit voltage (for take-over and re-ignition) will lead to a low cost ballast with small dimensions.

3.2 TARGET SPECIFICATIONS

3.2.1 Boost converter

\[ V_{in_{min}} = 9V \]
\[ V_{in_{cal}} = 12V, \text{ battery voltage used for calculations} \]
\[ V_{in_{nom}} = 13.2V, \text{ nominal battery voltage} \]
\[ V_{in_{max}} = 16V \]
\[ V_{out_{min}} = 1.2 \times V_{in}, \text{ duty-cycle control} \]
\[ V_{out_{max}} = 60V, \text{ maximum drain source voltage of the power mosfet} \]
\[ I_{out_{max}} = 2.6A, \text{ maximum electrode current of the lamp} \]
\[ P_{out_{run-up}} = 100W, \text{ increased lamp power for fast light run-up} \]
\[ P_{out_{steady-state}} = 35W, \text{ nominal lamp power in steady-state operation} \]

The Vedilis standard for the light run-up is:
- 25% of the nominal luminous flux within 1s after ignition
- 80% of the nominal luminous flux within 4s after ignition.
3.2.2 High-voltage circuit

Vin\textsubscript{nom} = 50V
Vin\textsubscript{max} = 60V, maximum output voltage of the boost converter
OCV\textsubscript{max} = 200V, maximum drain source voltage of the power mosfet
Vignitor\textsubscript{nom} = 800V, nominal breakdown voltage spark gap
Vignitor\textsubscript{max} = 920V, maximum breakdown voltage spark gap
Pin\textsubscript{continuous} = 5W, for ignition
Pin\textsubscript{peak} = 35W, for re-ignition

3.2.3 Commutator

Vin\textsubscript{max} = OCV = 200V, maximum drain source voltage of the power mosfets
Imax = 2.6A, maximum electrode current of the lamp
Imax = 10A, maximum take-over current
f\textsubscript{min} = 250Hz, maximum temperature variation of the electrodes

3.2.4 Ignitor

Vin\textsubscript{max} = 920V, maximum breakdown voltage spark gap
Vignition = 25kV, required ignition voltage for a hot-restrike
f\textsubscript{max} = 400Hz, maximum switching frequency spark gap
4 BOOST CONVERTER

The power converter transforms the power supplied from the battery to the power delivered to the commutator or the HV-circuit with minimum losses. The converter must be able to supply the variable power demand of the lamp during ignition, run-up and steady-state operation. An increased power or voltage demand has to be supplied fast, otherwise the lamp will extinguish. A boost topology converts to a higher voltage. The advantages of a boost converter are:
- an inductor (at the input) instead of a transformer,
- one of the basic topologies (1 mosfet, 1 diode, 1 inductor [20,22,23,24,25,26]),
- gate driving without level shifting.

An inductor has a lower price and volume (in comparison with a transformer) and it will allow for a continuous input current. On top of that stray inductance of the inductor is irrelevant for a good performance at high frequency. The electrical circuit is drawn in figure 4.1.

![Boost converter diagram](image)

Figure 4.1: Boost converter

4.1 STATE-SPACE AVERAGING

A method to analyse the static and dynamic performance of the boost converter is 'state-space averaging'. This method introduced by Middlebrook and Ćuk [30] gets more accurate the further the effective low-pass cut off frequency (determined by L, C, R and the duty-cycle) is below the switching frequency. However, this requirement is always satisfied because it is equivalent to the requirement for a small output voltage ripple (see section 4.5.1).
4.1.1 State-space equations

The boost converter operates in the continuous conduction mode, the inductor current does not drop to zero at any point in the switching cycle. The boost converter can be in two states; the mosfet is conducting (interval DT) or the diode is conducting (interval (1-D)T or D'T). The state of current commutation and the discontinuous conduction mode are left out of consideration. The electric circuits with parasitic resistances (R_{ds_{on}}=R_m, ESR_c=R_c) are depicted in figure 4.2 for both states.

![Figure 4.2: The two states of the boost converter](image)

State-space averaging for interval DT results in the following equations. The equations are derived with the aid of the Kirchhoff laws and superposition.

\[
\frac{di}{dt} = \left( \frac{R_i+R_m}{L} \right) i + \frac{V_i}{L} \tag{1}
\]

\[
\frac{dv}{dt} = \left[ \frac{1}{(R+R_c) C} \right] v \tag{2}
\]

\[ y = \frac{R}{R+R_c} v \tag{3} \]
State-space equations for interval $D'T$:

\[
\frac{di}{dt} = \left[ \frac{R_1 + R_d + R_c}{L} \right] i + \frac{R}{L(R + R_c)} v + \frac{v_i}{L} \tag{4}
\]

\[
\frac{dv}{dt} = \left[ \frac{R}{(R + R_c) C} \right] i + \frac{1}{(R + R_c) C} v \tag{5}
\]

\[
y = (R|R_c) i + \frac{R}{R + R_c} v \tag{6}
\]

4.1.2 State-space averaged equations

Averaging of the two states is obtained by summing:
- the equation for interval $DT$ (1, 2 or 3) multiplied by $D$
- and the equation for interval $D'T$ (4, 5 or 6 respectively) multiplied by $D'$.

The state-space averaged circuit equations are:

\[
\frac{di}{dt} = \left[ \frac{R_1 + DR_2 + D'R_d + D'R_c}{L} \right] i + \frac{D'R}{(R + R_c) C} v + \frac{v_i}{L} \tag{7}
\]

\[
\frac{dv}{dt} = \left[ \frac{D'R}{(R + R_c) C} \right] i + \frac{1}{(R + R_c) C} v \tag{8}
\]

\[
y = D'(R|R_c) i + \frac{R}{R + R_c} v \tag{9}
\]
4.1.3 Static model

The average inductor voltage and the average capacitor current are zero, hence it appears for the static model:

\[ 0 = \left[R_1 + D R_a + D' R_d + D' \frac{R}{R_c}\right] I - \frac{D' R}{R + R_c} V + V_I \quad (10) \]

\[ 0 = D' R I - V \quad I = \frac{V}{D' R} \quad (11) \]

\[ Y = D' \left(\frac{R}{R_c}\right) I + \left[\frac{R}{R + R_c}\right] V \quad (12) \]

Substitution of equation (11) in equation (10) results in an expression for the voltage gain:

\[ M = \frac{V}{V_I} = \frac{Y}{V_I} = \frac{1}{D'} \left[\frac{D' R}{D' R + R_1 + D R_a + D' R_d + D' D' \left(\frac{R}{R_c}\right)}\right] \quad (13) \]

The voltage gain is the ideal voltage gain \(1/D'\) corrected with a factor due to parasitic resistances. \(R_1\) is the most dominant parasitic resistance. The influence of \(R_m\) and \(R_d\) are determined by \(D\) and \(D'\) respectively. The equivalent series resistance \(R_c\) of the capacitor does hardly have any effect on the voltage gain.

4.1.4 Simplified state-space averaged equations

For obtaining the dynamic model the parasitic resistances are left out of consideration, this results in the following simplified state-space averaged equations:

\[ \frac{d i}{d t} = -\frac{D'}{L} \frac{V_I}{L} + \frac{V_I}{L} \quad (14) \]

\[ \frac{d v}{d t} = \frac{D' R}{C} \frac{V}{R_C} \quad (15) \]
Perturbation:
To analyse the response of the output voltage caused by a change in: lamp behaviour (for instance during run-up), battery supply voltage or duty-cycle (control circuit) the following perturbations are introduced:

\[ i = I + \dot{i} \]  
\[ v = V + \dot{v} \]  
\[ d = D + \dot{d} \quad d' = D' - \dot{d} \]  
\[ v_i = V_i + \dot{v}_i \]  
\[ r = R + \dot{r} \]  

Linearization:
Introduction of perturbations causes non-linearity. To obtain a small-signal dynamic model linearization has to be performed, according to a first order Taylor series.

\[ \frac{d\dot{i}}{dt} = -\frac{D'}{L} \dot{v} + \frac{1}{L} \dot{v}_i + \frac{V}{L} \dot{d} \]  
\[ \frac{d\dot{v}}{dt} = \frac{D'}{C} \dot{i} - \frac{1}{RC} \dot{v} - \frac{V_i}{D' RC} \dot{d} + \frac{V}{2 R^2 C} \dot{\phi} \]  

Laplace-transformation:

\[ s\dot{i}(s) - \dot{i}(0) = -\frac{D'}{L} \phi(s) + \frac{1}{L} \phi_i(s) + \frac{V_i}{D'} \phi(s) \]  
\[ s\dot{v}(s) - v(0) = \frac{D'}{C} \dot{i}(s) - \frac{1}{RC} \phi(s) - \frac{V_i}{D' RC} \dot{d}(s) + \frac{V_i}{D' 2 R^2 C} \phi(s) \]
Substitution:
Substitution from (23) in (24) with initial conditions set to zero

\[
sv(s) = \frac{D'}{sC} \left( -\frac{D'}{L} \hat{v}(s) + \frac{\hat{v}_1(s)}{L} - \frac{V_i}{D'L} \hat{d}(s) \right) - \frac{\hat{v}(s)}{RC} - \frac{V_i}{D'aRC} \hat{d}(s) + \frac{V_i}{D'a^2R^2C} \hat{p}(s) \]  

(25)

4.1.5 Dynamic model

Rearranging of equation (25) results in the desired expression for the dynamic performance of the boost converter:

\[
\hat{v}(s) = \frac{1}{LC + \frac{L}{RD} + \frac{L}{sD} + 1} \left[ \frac{1}{D'} \hat{v}_1(s) + \frac{V_i}{D'a} \left( 1 - \frac{SL}{D'^2R} \right) \hat{d}(s) + \frac{V_iLs}{D'^2a^2R^2} \hat{p}(s) \right] 
\]

(26)

The presented dynamic model is accurate for frequencies up to half the switching frequency. A representation in a block diagram of equation (26) with the introduction of an effective inductance is given in figure 4.3.

\[
\hat{L}_e = \frac{L}{D'^a} 
\]

(27)

Figure 4.3: Dynamic model of the boost converter

As expected an input voltage increase or decrease causes an output voltage increase or decrease respectively. A resistance increase or decrease causes a positive or a negative voltage dip respectively. The duty-cycle control has a right half-plane zero, this can cause complications when stabilizing the feedback.
control. The initial slope of the output response is negative for a positive input change, see figure 4.4.

![Figure 4.4: Transient response to an increased duty-cycle](image)

4.1.6 Transient response

The transfer function can be written in the standard form, according to Van de Vegte [37]:

$$F(s) = \frac{1}{L_{e}C s^{2} + \frac{L_{e} L}{R} s + 1} \frac{s^{2} + 2 \xi \omega_{n} s + \omega_{n}^{2}}{s^{2} + 2 \xi \omega_{n} s + \omega_{n}^{2}}$$

Hence it appears for the undamped natural frequency $\omega_{n}$ and the damping ratio:

$$\omega_{n} = \frac{1}{\sqrt{L_{e} C}} \quad \xi = \frac{1}{2R} \sqrt{\frac{L_{e}}{C}}$$

The speed of the response is increased by increasing the natural frequency. For a damping ratio equal to one the system is critically damped, for $d.r. < 1$ the system is underdamped and overshoot over the steady-state response appears. A boost converter with a fast transient response is obtained when the following design criteria are satisfied:

- inductor with low inductance
- output capacitor with low capacitance
- low duty-cycle: high input voltage, since the output voltage is set by the lamp.

Furthermore, the resistance of the load determines whether or not the system is damped.
4.2 SWITCHING FREQUENCY AND OPERATION MODE

To design a ballast with small components for achieving a low price, a small volume and a fast transient response, the switching frequency has to be high. The disadvantages of a high switching frequency are the increased switching losses in the semiconductors. The switching frequency is set to 500kHz. This is an optimum in case a low cost automotive ballast is required [12]. Because of this high switching frequency a fast transient response can be obtained (see section 4.1.6) and a re-ignition coil is not necessary.

The steady-state operation mode chosen is at the boundary of continuous conduction mode (CCM) and discontinuous conduction mode (DCM). The advantages of this operation mode are:
- an inductor with low inductance for a fast transient response
- on-switching of the power mosfet without dissipation
- no reverse recovery losses of the diode

disadvantage:
- high ripple current: \( I_{\text{rms}} = \frac{2}{\sqrt{3}} \times I_{\text{av}} \approx 1.15 \times I_{\text{av}} \).

The conditions for boundary operation are a battery voltage of 12V and a lamp voltage of 50V. The power supplied by the battery is 35W lamp power corrected with an estimated converter efficiency of 90%. This results in a duty-cycle of 76% (ideal voltage gain according to equation (13)) and an average input current of 3.25A. The inductor current is shown in figure 4.5.

![Figure 4.5: Desired inductor current, 35W operation](image-url)
The desired inductor value is:

\[ L = V \frac{dt}{di} = 12 \frac{1.52 \mu s}{6.5 A} = 2.81 \mu H \]  

(30)

The maximum input current occurs at an input voltage of 9V during run-up. The duty-cycle for 50V output voltage is 0.82 (equation 13), the estimated converter efficiency for run-up is 80%:

\[ I_{in, max} = \frac{P_{run-up}}{V_{in, min} \cdot \eta} = \frac{50V \cdot 2.6A}{9V \cdot 0.8} = 18.1A \]  

(31)

The peak value of the inductor current at this maximum input current (DT is 1.64μs) is:

\[ I_{L, max} = I_{in, max} + \frac{9V \cdot 1.64 \mu s}{2.81 \mu H} = 20.7A \]  

(32)

4.3 INDUCTOR DESIGN

4.3.1 Magnetic core

The magnetic core is made of soft ferrite. The available ferrites contain compositions of manganese and zinc and have good magnetic properties below the Curie-temperature. The ferrite choice is a compromise of price and specific power loss with the switching frequency as main parameter. The power loss as function of temperature for the ferrites considered is given in figure 4.6 (Soft Ferrites [39]).
The ferrite 3C80 happens to be the most commonly used and the most inexpensive material but unfortunately also the ferrite with the highest power loss. The saturation flux density at 25°C is 420mT, the saturation flux density at 100°C is 330mT. A saturation flux density of 350mT is chosen for calculations. In case the magnetic flux is homogeneously distributed over the effective area, the maximum magnetic flux for the cores considered is:

- U10: \( A_e = 8.6 \text{mm}^2 \), \( \Phi_{\text{max}} = 3.0 \mu \text{Vs} \)
- U11: \( A_e = 17.6 \text{mm}^2 \), \( \Phi_{\text{max}} = 6.2 \mu \text{Vs} \)
- U15: \( A_e = 32.3 \text{mm}^2 \), \( \Phi_{\text{max}} = 11.3 \mu \text{Vs} \)

### 4.3.2 Number of turns

The minimum number of turns can be calculated with Ampère's law:

\[
\oint Hdl = \iint (J + \frac{d\Phi}{dt}) \cdot ds
\]  

(33)

The \( d\Phi/dt \) displacement current density differs from zero only at current commutation. This small displacement current density can
be neglected in comparison with the high current density. Hence it appears for the static Ampère's law:

$$\oint H dI = NI \quad B = \mu H$$  \hspace{1cm} (34)$$

$$\frac{B_f}{\mu _f \mu _0} I_f + \frac{B_i}{\mu _i \mu _0} I_i = NI$$  \hspace{1cm} (35)$$

If the air gap is small compared to the dimensions of the core and $\mu _i >> 1$, the following approximation for the magnetic flux inside the core is valid:

$$\Phi = A_s B = \frac{A_s \mu _0 N I}{I_a}$$  \hspace{1cm} (36)$$

According to Faraday's law, the formula for the inductance is:

$$\oint E dI = - \frac{d}{dt} \int B \cdot dS \quad v_i = N \frac{d\Phi}{dt} = L \frac{di}{dt}$$  \hspace{1cm} (37)$$

$$L = \frac{\mu _0 A_s N^2}{I_a}$$  \hspace{1cm} (38)$$

The number of turns has to satisfy the following condition:

$$N \geq \frac{L_i I_{max}}{\Phi_{max}} = \frac{58.2 \mu V S}{\Phi_{max}}$$  \hspace{1cm} (39)$$

U10: \hspace{0.5cm} N_{min} = 20 \hspace{0.5cm} l_a > 1.54mm \\
U11: \hspace{0.5cm} N_{min} = 10 \hspace{0.5cm} l_a > 0.78mm \\
U15: \hspace{0.5cm} N_{min} = 6 \hspace{0.5cm} l_a > 0.52mm \\

The core selected is the U11 core of 3C80 ferrite. This core is an optimum between the number of turns and the ferrite volume. The inductor can also be designed with a realistic air gap.
4.3.3 Air gap

The core consists of two similar U-parts with the air gap in between, the total air gap is twice the distance of the U-parts. The air gap is filled with paper with a thickness of 0.5mm (the air gap is in fact a paper gap). The total geometric air gap is 1mm. The 'magnetic' air gap is smaller due to the expanding effect of magnetic field lines in the air gap region, see figure 4.7. An advantage of a relatively large air gap is the reduced influence of a lateral displacement of the core parts on the inductance.

![Figure 4.7: Expanding magnetic field lines](image)

The calculated inductance of the U11 inductor with an air gap of 1mm and 10 turns is $2.21\mu\text{H}$ (equation 38). The real inductance will be higher due to a smaller 'magnetic' air gap and the presence of stray inductance.

4.3.4 Winding

The resistance of the winding increases with the frequency. This is caused by skin and proximity effects. An alternating magnetic field is driven out of the conductor by a nonhomogeneous current distribution inside the conductor (Casimir and Ubbink [38]). For a high frequency current flowing through a conductor with a high permeability the highest current density will be near the boundary of conductor and air, the skin of the conductor.
The boost converter has a triangle wave inductor current. The switching frequency is 500kHz (3rd harmonic at 1.5MHz), solid wire would result in a high copper resistance at 500kHz. To minimize skin and proximity effects Litz-wire has been taken. The Litz-wire used is a bunch of 60 cyclic twisted wires of 0.1mm diameter each.

4.3.5 Inductor measurement

The inductance and the resistance are measured as function of the frequency with a Hewlett Packard impedance/gain-phase analyzer 4194A, see figure 4.8. The measurement voltage is reduced to an allowable minimum to cancel core losses.

The measured inductance is 3.3μH. The measured resistance for direct current is 20mΩ and 94mΩ for a harmonic current with a frequency of 500kHz. The spectrum of the inductor current contains a DC component (main component), a 500kHz component and higher harmonic components. Calculations will be performed with a simplified frequency independent resistance of 50mΩ.
4.4 SEMICONDUCTOR SELECTION

4.4.1 Power mosfet selection

The selection of the power mosfet is based on price, maximum voltage and current, $R_{ds_{on}}$ and gate capacitance. The gate capacitance determines the switching speed and therefore the power dissipation in the mosfet during switching. The $R_{ds_{on}}$ determines the power dissipation in the mosfet during conduction. The power mosfets considered are (values for 25°C case temperature):

<table>
<thead>
<tr>
<th>Vds_{max} (V)</th>
<th>Ids_{max} (A)</th>
<th>Rds_{on} (Ω)</th>
<th>Ciss (pF)</th>
<th>50k</th>
<th>500k</th>
</tr>
</thead>
<tbody>
<tr>
<td>SGS-Thomson:</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>STP50N06</td>
<td>60</td>
<td>50</td>
<td>0.028</td>
<td>1700</td>
<td>DM 1.20</td>
</tr>
<tr>
<td>International Rectifier:</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>IRFZ44</td>
<td>60</td>
<td>50</td>
<td>0.028</td>
<td>1900</td>
<td>$0.72$</td>
</tr>
<tr>
<td>IRFZ48</td>
<td>60</td>
<td>50</td>
<td>0.018</td>
<td>2400</td>
<td>$1.03$</td>
</tr>
<tr>
<td>Philips:</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>BUK456-60A</td>
<td>60</td>
<td>52</td>
<td>0.028</td>
<td>1500</td>
<td>DFL 1.71</td>
</tr>
<tr>
<td>BUK456-60B</td>
<td>60</td>
<td>51</td>
<td>0.030</td>
<td>1500</td>
<td>DFL 1.65</td>
</tr>
<tr>
<td>Motorola:</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>MTP50N06E</td>
<td>60</td>
<td>50</td>
<td>0.025</td>
<td>2500</td>
<td>$1.28$</td>
</tr>
</tbody>
</table>

According to the present exchange rates the SGS-Thomson power mosfet is the least expensive and is therefore selected. The STP50N06 also has a low gate capacitance. The gate charge and capacitance graphs are given in figure 4.9.

Figure 4.9: Gate charge and capacitance of the STP50N06
4.4.2 Power diode selection

The current commutation from the power diode to the mosfet has to occur with minimum losses. During run-up the current is high and a fast reverse recovery diode is required then. For obtaining a high efficiency the forward voltage has to be low. The ultra fast recovery rectifier diode Philips BYW29 satisfies both the requirements.

![Graph showing forward voltage and stored charge of the BYW29 diode.](image)

**Figure 4.10: Forward voltage and stored charge of the BYW29**

Reverse recovery current is caused by diffusion of minority carriers. This stored charge is determined by the forward current just before the current commutation and is removed by diffusion and recombination. The time needed to remove the stored charge, the reverse recovery time, decreases with a shorter minority carrier lifetime. Due to recombination of minority carriers the stored charge decreases with a slower switching speed, according to figure 4.10.

BYW29-100 (nominal values): $V_g=100\text{V}$, $V_f=0.8\text{V}$, $I_f=8\text{A}$, $t_{rr}=25\text{ns}$.
4.5 CAPACITOR SELECTION

4.5.1 Boost capacitor

The capacitor at the output of the boost converter has to have a low capacitance for a fast transient response. The minimum capacitor value is determined by a requirement for a maximum ripple at which acoustic resonances will not occur. The maximum allowed top-top value of the ripple is 10% of the average output voltage. The capacitor value is calculated as follows:

\[ 47.5V = 52.5Ve^{-\frac{1.52\mu F}{RC}} \]  \hspace{1cm} (40)

\[ R = \frac{(50V)^2}{35W} = 71.4\Omega \quad C = 213nF \]  \hspace{1cm} (41)

A Philips MKT 370 polyester film capacitor of 220nF/63V is used. The capacitance and the equivalent series resistance (ESR) are measured as function of the frequency, see figure 4.11.

Figure 4.11: Capacitance and ESR of the boost capacitor

4.5.2 Input capacitor

The input current ripple can be reduced by placing a capacitor close to the inductor and the mosfet of the boost converter. This
capacitor causes a battery current with a smaller ripple and as a result a lower electromagnetic interference.

According to figure 4.5 the charge consumed by the boost converter per switching cycle is 6.5\(\mu\)C. For a ripple free battery current and a converter operating at the boundary of CCM and DCM the input capacitor is charged and discharged each cycle with half of the charge consumed, 3.25\(\mu\)C. The allowable voltage dip caused by discharging is 1V (a chosen value). So the desired input capacitance equals 3.25\(\mu\)F. Three 1\(\mu\)F/63V multilayer ceramic capacitors (small dimensions) of Siemens are used, total input capacitance is 3\(\mu\)F. The ceramic material is X7R, the capacitance and the series resistance of one capacitor are measured as function of the frequency.

![Figure 4.12: Capacitance and ESR of one input capacitor](image)

Both the input and output capacitor have to be placed very close to the inductor, power mosfet and power diode. A lot of attention has to be paid to the lay-out of the boost converter. The power mosfet and the diode have to be mounted on a heatsink because of their heat generation. The thermal resistance from junction to case is for the mosfet 1°C/W and for the diode 2.7°C/W (for free air operation the thermal resistance from junction to ambient for both semiconductors is 60°C/W).
4.6 POWER LOSSES CALCULATION

The power losses of the boost converter are calculated for 35W operation, 12V input voltage, 50V lamp voltage and 25°C ambient temperature. The electric schematic including the parasitic losses is given in figure 4.13.

![Boost converter schematic](image)

**Figure 4.13: Boost converter with parasitics**

The assumed input power is the nominal output power corrected with an estimated converter efficiency (90%), this results in 38.9W input power. The inductor current is given in figure 4.14.

![Inductor current graph](image)

**Figure 4.14: Calculated inductor current, 35W operation**

4.6.1 Inductor losses

The AC component of the inductor current has a peak value of 2.76A, the measured inductance is 3.3μH. The amplitude of the
alternating magnetic swing is:

$$E = \frac{\Phi}{A_e} = \frac{L}{N. A_e} = 52 \text{mT}$$

(42)

A magnetic swing of 52 mT around 0 mT is assumed to have an equal core loss as an magnetic swing of 52 mT around 61 mT (3.24 A DC component of the current). The triangle wave excitation of the magnetic core is assumed to have an equal core loss as a harmonic excitation with the same amplitude. Fourier transformation of the triangle wave excitation and calculation of the core loss per frequency component is not correct because of the non-linearity and hysteresis of the ferrite core. The specific power loss for a 3C80 core with a pure harmonic excitation (f=500 kHz, B=50 mT, T=25°C) is 490 kW/m³, according to figure 4.6. The core loss (2 parts) can be approximated with the assumptions mentioned:

$$P_{fe} = 490 \times 10^{-6} \frac{W}{mm^3} \times 17.6 mm^2 \times 22.9 mm \times 2 = 0.39 W$$

(43)

The copper resistance is frequency dependent. For calculation of the power dissipated the equivalent constant resistance is 50 mΩ.

Intermezzo: calculation of the rms-value of the inductor current

$$I_{1,rms} = \frac{1}{T} \int_{T} I_1^2(t) \, dt$$

(44)

$$I_{1,rms}^2 = \frac{1}{1.52 \mu \text{s}} \int_{0.13 \mu \text{s}}^{1.65 \mu \text{s}} \left[ 3.63 \frac{A}{\mu \text{s}} \times t(\mu \text{s}) \right]^2 \, dt = 13.03 A^2$$

(45)

$$I_{1,rms} = 3.61 A$$

(46)

end of intermezzo.

Hence it appears for the copper losses:

$$P_{cu} = I_{1,rms}^2 \cdot R_{equ} = 0.65 W$$

(47)
4.6.2 Mosfet losses

The power mosfet is switched on and off by charging and discharging respectively of the gate. The turn-on and the turn-off times are determined by the source and sink current of the gate driver. The gate driver used in the experimental set-up is a IR2110, this is a regenerative mos-gate driver capable of 2A source and sink current (Clemente [35]).

The gate charge at the threshold voltage (Vgs=Vth=3V) is 7nC. The gate charge required for charging Cgd is 26nC, flat section in figure 4.9. Difference in charge between Vd=40V and Vd=50V can be neglected; the capacitance Cgd decreases fast with an increasing Vd. The total gate charge required for switching on a mosfet (Ids=Ii=0.48A, Vds=Ids*Rds(on)) is 33nC. Due to the presence of parasitic inductance and resistance in the MGD mosfet connection, the average source and the average sink current are estimated at 1A (a smaller value than the maximum MGD drive current). This results in a turn-on time of 26ns. Switching-off the mosfet occurs at 6A drain current. Because of the high forward transconductance (gfs=22A/V) the turn-off time will be equal to the turn-on time.

Due to the high transconductance almost the entire turn-on time or turn-off time is necessary for charging or discharging the drain gate capacitance. The power dissipated for switching on the mosfet can be calculated as follows (Mohan [17], Clemente [36]):

\[
P_{\text{turn-on}} = f_{\text{switch}} \cdot \frac{V_{ds}}{2} \cdot I_{ds} \cdot t_{on}
\]

\[
P_{\text{turn-on}} = 500 \text{kHz} \cdot \frac{50V}{2} \cdot 0.48A \cdot 26ns = 0.16W
\]

Dissipated power for switching off:

\[
P_{\text{turn-off}} = 500 \text{kHz} \cdot \frac{50V}{2} \cdot 6.0A \cdot 26ns = 1.95W
\]

The drain source capacitance is charged when switching off the
mosfet. The charge current is a part of the current flowing through the inductor. The 6A inductor current equals the mosfet drain current which is divided in a channel current (main part 90%) and a displacement current caused by Cds. Therefore Cds is charged by a current source and charging occurs without dissipation. Switching the mosfet on causes discharging of Cds. The stored energy is dissipated in the mosfet. The voltage dependent depletion capacitance is approximated by a constant capacitance of 400pF. The power dissipated equals:

\[ P_{Cds} = f_{switch} \cdot \frac{1}{2} \cdot C_{ds} \cdot V_{ds}^2 = 0.25W \]  \hspace{1cm} (51)

The power losses caused by the \( R_{ds_{on}} \) during conduction are:

\[ P_{Rds_{on}} = R_{ds_{on}} \cdot I_{1,\text{rms}}^2 \cdot D = 0.28W \]  \hspace{1cm} (52)

The total power dissipated in the mosfet is 2.64W.

4.6.3 Gate driver losses

The output stage of the mos-gate driver IC is a half bridge with NMOS transistors with a high W/L ratio. The MGD is supplied with an auxiliary voltage of 12V. The gate charge of the power mosfet at 12V gate voltage is 50nc. This gate charge is supplied by the driver. The power dissipated in the mos-gate driver is:

\[ P_{MGD} = f_{switch} \cdot V_A \cdot Q_g = 0.3W \]  \hspace{1cm} (53)

4.6.4 Diode losses

The power losses caused by the forward voltage (0.7V) are:

\[ P_F = V_F \cdot I_{1,av} \cdot (1-D) = 0.55W \]  \hspace{1cm} (54)

Power losses caused by the 12mΩ series resistance (figure 4.10):

\[ P_{Rd} = R_r \cdot I_{1,\text{rms}}^2 \cdot (1-D) = 0.04W \]  \hspace{1cm} (55)
Calculation of the reverse recovery losses, according to the voltage and current waveforms depicted in figure 4.15.

\[ t \approx 100\% \]

Figure 4.15: Reverse recovery

The stored charge is approximated by extrapolation in figure 4.10 \((I_r=0.5A, \; -\frac{dI_r}{dt}=100A/\mu s)\), \(Q_s\) is 13nC.

\[
P_{rr} = f_{\text{switch}} \cdot \frac{V_R}{2} \cdot \frac{Q_s}{2} = 0.08W
\]  
(56)

The total power dissipated in the diode is 0.67W.

4.6.5 Capacitor losses

In the (1-D) interval the discharge current of the output capacitor is 0.8A. In the D interval the peak charge current is 5.1A, this charge current drops to zero at the end of the D interval. The power losses in the ESR of the boost capacitor are:

\[
P_{c_{\text{boost}}} = \left[ (0.8A)^2D + \left( \frac{5.1A}{2} \times 1.15 \right)^2D \right] \cdot 45\Omega = 0.12W
\]  
(57)
The amplitude of the alternating triangle wave current in the input capacitor is 2.76A. The power losses in the ESR of the input capacitor are:

\[ P_{\text{Cin}} = ESR \cdot I_{\text{rms}}^2 = 14 \text{m\Omega} \cdot \left( \frac{2.76 \text{A}}{2 \cdot 1.15} \right)^2 = 0.04 \text{W} \]  

4.6.6 Power balance and efficiency

The power supplied by the battery is assumed to be 38.9W, power losses per component are (gate driver losses are not included):

- Inductor: 1.04W
- Mosfet: 2.64W
- Diode: 0.67W
- Boost capacitor: 0.12W
- Input capacitor: 0.04W

The total power losses are 4.51W, the main part is dissipated in the mosfet. This is convenient because the heat generated through this power dissipation is easy to transport by a heatsink (the ballast case). Power delivery to the commutator is 34.4W.

The calculated efficiency at 25°C ambient temperature is:

\[ \eta = \frac{P_{\text{out}}}{P_{\text{in}}} = 88\% \]  

4.7 PERFORMANCE

The operating area of the boost converter is depicted in an output voltage versus output current graphic, figure 4.16. The absolute maximum ratings of the output voltage (V_{ds_{\text{max}}} of the mosfet) and output current (maximum electrode current) are indicated.
The period time of the undamped natural frequency is an indication for the transient response. For a lamp just after ignition (lamp voltage immediately after breakdown is 25V) the undamped period time is, according to equation (29):

$$T_n = 2\pi \sqrt{\frac{LC}{D'}}$$

For a lamp in steady-state operation this period time is:

$$T_n(V_{lamp} = 25V) = 10.3 \mu s$$

(60)

$$T_n(V_{lamp} = 50V) = 22.3 \mu s$$

(61)

The transient response time can become shorter when a well matched feedback control is implemented.

4.8 MEASUREMENTS

4.8.1 Current and voltage measurements

The input voltage is measured with a Hewlett Packard 3403C true rms voltmeter, the output voltage is measured with a Tektronix 2430A digital oscilloscope and a voltage probe. The current is measured with a Tektronix A6302 current probe and a current probe amplifier AM503. The inductor current and the output voltage are measured for 12V battery voltage and 35W power delivery to a resistor.
The output voltage satisfies the 10% requirement, the inductor current is in accordance with calculation (figure 4.14). The battery current is measured under similar conditions.

The battery current equals the average inductor current. The DC battery current will not cause electromagnetic interference.
4.8.2 Power measurement

Power analyzers are not used because of their limited bandwidth. The advantage of voltage measurement by oscilloscope is the fact that the output voltage can be viewed. Mismeasurement by a device with a bandwidth that is too small can therefore be avoided. The oscilloscope has a relatively large inaccuracy of 5%. Two MKT film capacitors of $6.8\mu F$ each are placed at the output of the boost converter to reduce the output voltage ripple.

The measured efficiency as function of the input voltage is:
for 50V output voltage and 35W power delivered to a resistor (gate driver losses are not included)

<table>
<thead>
<tr>
<th>Vin (V)</th>
<th>D</th>
<th>mode</th>
<th>n (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>9</td>
<td>0.82</td>
<td>CCM</td>
<td>79</td>
</tr>
<tr>
<td>12</td>
<td>0.76</td>
<td>CCM</td>
<td>85</td>
</tr>
<tr>
<td>13.2</td>
<td>0.74</td>
<td>CCM</td>
<td>87</td>
</tr>
<tr>
<td>16</td>
<td>0.55</td>
<td>DCM</td>
<td>89</td>
</tr>
</tbody>
</table>

The measured efficiency as function of the output voltage is:
for 12V input voltage and 35W power delivery (all in CCM)

<table>
<thead>
<tr>
<th>Vout (V)</th>
<th>D</th>
<th>n (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>20</td>
<td>0.43</td>
<td>92</td>
</tr>
<tr>
<td>30</td>
<td>0.61</td>
<td>90</td>
</tr>
<tr>
<td>40</td>
<td>0.70</td>
<td>88</td>
</tr>
<tr>
<td>50</td>
<td>0.76</td>
<td>85</td>
</tr>
<tr>
<td>60</td>
<td>0.80</td>
<td>83</td>
</tr>
</tbody>
</table>

The efficiency as function of the output power is:
for 12V input voltage and 50V output voltage

<table>
<thead>
<tr>
<th>Pout (W)</th>
<th>n (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>35</td>
<td>85</td>
</tr>
<tr>
<td>50</td>
<td>85</td>
</tr>
<tr>
<td>70</td>
<td>85</td>
</tr>
<tr>
<td>100</td>
<td>81</td>
</tr>
</tbody>
</table>
The efficiency is nearly constant over a wide range of the output power. The efficiency decreases significantly with an increased output voltage or with a decreased input voltage. The ratio of output voltage and input voltage has to be low for obtaining a high efficiency!

4.9 VERIFICATION OF CALCULATION AND MEASUREMENT

Power loss calculation:
The measured efficiency for \( V_{in} = 12V \), \( V_{out} = 50V \) and \( P_{out} = 35W \) is 85\%, the calculated efficiency is 88\%. The calculated efficiency comes close to the measured efficiency. The presented power losses calculation (chapter 4.6) is correct.

Static model:
The static model (equation 13) derived from the state-space averaged equations is in accordance with the measurements and is therefore correct.

Dynamic model:
The dynamic model (equation 26) is verified with a Spice computer simulation. Input is the electric schematic of the boost converter with parasitics. The duty-cycle is fixed at 0.76, the battery voltage is 12V. The load resistance is suddenly decreased from 7Ω (35W power delivery) to 25Ω (100W power delivery). With this fixed duty-cycle the boost converter needs 20μs for achieving a steady-state output voltage. This simulated transient response time, with fixed duty-cycle, is in accordance with the period time of the undamped natural frequency as given by equation (60) and (61).
4.10 LIMITED VOLTAGE GAIN FOR POWER CONVERSION

The inexpensive boost converter is not suitable for the present 85V MHS lamp with 12V battery voltage. Power conversion with this high voltage gain is not possible because:

- The efficiency decreases considerably with a 85V lamp and 12V battery voltage. The efficiency for 35W operation is below 75%, based on extrapolation of the measurement results in section 4.8.2. This will cause serious problems with heat transport, especially during the run-up with 100W output power.

- A power mosfet with a higher maximum drain source voltage (100V) is needed (higher price). The higher input capacitance results in an increase of the mosfet switching losses. Power dissipation in the mosfet for 35W operation will be about 5W, section 4.6.2.

- The high duty-cycle of 86% causes control and stability problems at 500kHz switching frequency (see section 4.1.5, 4.1.6 and 4.2). The switching frequency has to be reduced. This results in more expensive and larger components.

In the worst case situation with a 9V battery voltage and 102V lamp voltage (duty-cycle is 91%) a good performance is even not possible.

Power conversion can be realized in an inexpensive way in case the ratio of lamp voltage and battery voltage is low.
5 TAPPED-INDUCTOR BOOST CONVERTER

The following topologies are considered for generating the ignitor voltage and the open circuit voltage:
- flyback,
- boost,
- diode capacitor voltage multiplier.

The ignitor capacitor has to be charged up to 800V, the open circuit voltage is 200V. This two different voltages can be obtained by:
- a resistance and/or capacitance divisor,
- or by inductor tapping.

Investigation of the possibilities mentioned has resulted in a tapped-inductor boost converter [27,31]. This converter is the best solution to achieve a low cost circuit for generating the desired voltages. The electrical circuit of the tapped-inductor boost converter with dual output is given in the figure below.

![Figure 5.1: Tapped-inductor boost converter with dual output](image)

5.1 GENERAL ANALYSIS

The HV-circuit charges the take-over capacitor and the ignitor capacitor. The HV-circuit does not operate in a steady-state mode similar to the boost converter. In spite of this a derivation of the steady-state voltage gain for a single output (figure 5.2) in continuous conduction mode of the primary inductance is presented here.
Interval DT, the mosfet is conducting:
The current $i$ is the current flowing through the primary
inductance during interval DT.

$$\frac{di}{dt} = \frac{V_i}{L_1}$$  \hspace{1cm} (62)

Interval D'T, the diode is conducting:
The current $i'$ is the current flowing through the primary
inductance during interval D'T.

$$\frac{di'}{dt} = \frac{V_i - V_o}{L_1 N_1^2 \left( \frac{N_1 + N_2}{N_1} \right)^2}$$  \hspace{1cm} (63)

According to equation (34) results for the current in the primary
inductance just before ($dT^-$) and just after ($dT^+$) the current
commutation from mosfet to diode:

$$i(dT^-) = i'(dT^+) * (N_1 + N_2)$$  \hspace{1cm} (64)

The current in the primary inductance with $N_1 = N_2$ is depicted in
figure 5.3.
Substitution of equation (64) in equation (63) results with equation (62) in the averaged circuit equation for the inductor current $i$:

$$\frac{dl}{dt} = D \frac{V_i}{L_1} + D' \frac{V_i - V_o}{L_1 \left( \frac{N_1 + N_2}{N_1} \right)}$$  \hspace{1cm} (65)$$

The voltage gain is obtained by setting equation (65) to zero. Hence it appears for the static voltage gain:

$$M = \frac{V_o}{V_i} = \frac{1 + D \frac{N_2}{N_1}}{D'}$$  \hspace{1cm} (66)$$

In comparison with a normal boost converter the tapped-inductor boost converter has a higher voltage gain with the same duty-cycle. The voltage stress of the switch is reduced, the 800V ignitor voltage can be generated with a 200V mosfet. However the diode reverse voltage ($V_r$) is increased:

$$V_r = V_o + \frac{N_2}{N_1} V_i$$  \hspace{1cm} (67)$$

Although the input current does not drop to zero the current shows a discontinuity at the current commutation, see figure 5.3. This discontinuous inductor current can cause EMI rather than for instance the 500kHz inductor current of the boost converter.

5.2 TURNS RATIO

The magnetic flux inside a transformer is canalized. The voltage per turn is therefore equal for both the windings of the transformer. Breakdown of the spark gap inside the ignitor shows a statistical behaviour. Breakdown occurs by a Townsend mechanism [6] and has a tolerance of 15%.

$$680V < V_{bd_{gg}} < 920V$$  \hspace{1cm} (68)$$

In case the spark gap breaks down at maximum voltage the open
circuit voltage (OCV) may not exceed the 200V limit. With a boost voltage of 50V, a spark gap breakdown at 920V and a safe OCV of 195V appears for the turns ratio of the transformer:

$$\frac{N_2}{N_1} = \frac{920V - 195V}{195V - 50V} = 5$$  \hspace{1cm} (69)

If the spark gap breaks down at the nominal voltage the OCV will be 175V. At the minimum spark gap breakdown voltage (680V) is the OCV 148V. Successful take-over is doubtful then and a second ignitor pulse could be necessary.

5.3 TRANSFORMER DESIGN

The transformer has to have a good couple factor to reduce the effects of stray inductance, especially at 500kHz. Furthermore, the transformer has to be inexpensive and small in size. The E13 core is the smallest E-core available. A transformer containing two E13 cores satisfies both requirements and is selected therefore.

E13: $Ae=13.0\text{mm}^2$, $\phi_{\text{max}} = 4.6\mu\text{Vs}$

The ferrite chosen is 3C80 [39], the relative incremental permeability is 2000. The incremental permeability is observed when an alternating magnetic field is superimposed on a static bias field $H_{\text{DC}}$:

$$\mu_a = \frac{1}{\mu_0} \left[ \frac{\Delta B}{\Delta H} \right]_{H_{\text{DC}}}$$  \hspace{1cm} (70)

The effective magnetic length of a transformer built with two E13 cores is 29.6mm. The coilformer has two sections, one for the primary 200V winding and the other for the secondary 800V winding. This convenient separation results in a better insulation and a lower parasitic capacitance.

Rough calculation of the desired primary inductance:

Ignition is supposed to consume 5W power (100 ignition and take-
over pulses of 25mJ each per second). With a boost voltage of 50V this results in an average primary inductor current of 0.1A. The current in the D'T interval is left out of consideration for this rough calculation. This assumption is valid because of the relatively high turns ratio. With a duty-cycle of 25% and 500kHz switching frequency appears for the desired primary inductance:

\[ L_1 = \frac{\int v(t) \, dt}{i(t_1) - i(t_0)} \]

The top value of the primary inductor current for this supposed condition is 0.8A. With this 5W power consumption the minimum number of primary turns, according to equation (39) equals 6. A transformer with this minimum number of turns would have an unrealistic small air gap (2*9.5\mu m). A paper with a thickness of 0.05mm is used for the air gap. The number of primary turns is set to 14. According to equations (35) and (37) the primary inductance is:

\[ L_1 = \frac{\mu_0 A_s N_1^2}{l_1 + \frac{I_s}{\mu_0}} = 27.9 \mu H \]

The primary inductor current is allowed to rise to 2.3A (equation 39) before saturation occurs. The winding is a solid wire with 0.25mm diameter. The secondary winding has 70 turns (equation 69). Almost the entire winding space in the coilformer is filled.

The primary and secondary inductance are measured at 500kHz with minimum measuring voltage to cancel core losses.

L1 (sec open) = 32.1\mu H
L2 (prim open) = 810\mu H

A method for determination of the mutual inductance \(L_m\) is by measurement of the inductances \(L_a\) and \(L_b\) of the two possible series connections of the primary and secondary windings [27]:
\[ L_A = L_1 + L_2 + 2L_m = 1134 \mu H \] 
(73)

\[ L_B = L_1 + L_2 - 2L_m = 538 \mu H \] 
(74)

\[ L_m = 149 \mu H \] 
(75)

The coefficient of coupling \( k \) equals:

\[ k = \frac{L_m}{\sqrt{L_1 L_2}} = 0.92 \] 
(76)

5.4 SEMICONDUCTOR SELECTION

The mosfet used is the IRF610 from International Rectifier, this is an inexpensive one with a small chip area:

IRF610: \( V_{ds_{\text{max}}} = 200V \), \( I_{ds_{\text{max}}} = 3.3A \), \( R_{ds_{\text{on}}} = 1.5\Omega \), \( C_{iss} = 140\text{pF} \).

The \( I_{ds_{\text{max}}} \) at 100°C is 2.1A, the maximum primary inductor current is 2.3A (saturation). So these current ratings are matched to each other.

The diode D1 has to have a fast recovery, the BYV26B is selected.

BYV26B: \( V_R = 400V \), \( V_f = 2.5V \), \( I_f = 1A \), \( t_{rr} = 30\text{ns} \).

The diode D2 has to be capable of reversing a voltage of 1170V, equation (67). A series connection of two BYV26D diodes is used.

BYV26D: \( V_R = 800V \), \( V_f = 2.5V \), \( I_f = 1A \), \( t_{rr} = 75\text{ns} \).

The diode between the boost converter and the commutator has to reverse the open circuit voltage. In run-up and stead-state operation it has to have a low forward voltage and it must conduct the maximum lamp current (2.6A). The BYV27-200 is selected.

BYV27-200: \( V_R = 200V \), \( V_f = 1.07V \), \( I_f = 3A \), \( t_{rr} = 25\text{ns} \).

The power dissipation during steady-state operation is:

\[ P_{\text{diode}} = V_f \cdot I_{\text{lamp}} = 0.75W \] 
(77)
The diode between the HV-circuit and the commutator has to conduct the peak take-over current. The diode must prevent charging from the take-over capacitor directly by the boost converter, since this charging would result in a slow transient response of the boost converter. The BYV27-100 is selected.

BYV27-100: \( V_R = 100\text{V}, \ V_F = 1.07\text{V}, \ I_T = 3\text{A}, \ t_{tr} = 25\text{ns}. \)

5.5 TAKE-OVER CAPACITOR

The take-over current is the current flowing through the lamp immediately after breakdown (the lamp voltage is then 25V). This current is obtained from the discharge of the take-over capacitor. At the moment the capacitor's voltage equals the boost voltage, the boost converter has to supply the lamp current \([8]\). The boost converter with fixed duty-cycle needs 10\(\mu\text{s}\) (equation 60) to fulfil this increased power demand. The maximum take-over current is set to 10A (based on experience). The minimum capacitance is:

\[
C_{\text{take-over}} = \frac{10\text{A} \times 10\mu\text{s}}{175\text{V} - 25\text{V}} = 0.67\mu\text{F}
\]  

The capacitor chosen is a Siemens MKP polypropylene film capacitor of 1\(\mu\text{F}/250\text{V}\). The capacitance and ESR are measured, see figure 5.4.

![Figure 5.4: Capacitance and ESR of the take-over capacitor](image-url)
In case the impedance of the series ignitor (resistance and inductance) is very low the discharge of the take-over capacitor has to be controlled with a current limitation. The limitation of the discharge current (10A) can be incorporated in the commutator. Furthermore, the take-over current may not drop to zero before the moment that the converter is able to supply lamp power.

5.6 PERFORMANCE

All the components have to be placed very close to each other. The distance between the output capacitor of the boost converter and the transformer and mosfet of the HV-circuit has to be reduced to a minimum to avoid unnecessary increase of stray inductance.

During ignition of the lamp the take-over capacitor discharges fast; the high di/dt value of this current induces unwanted voltages. To avoid this induction the wires where the discharge current is flowing through have to be placed very close to each other to reduce an external dφ/dt field.

The 800V wire from the transformer to the ignitor has a diameter of 0.25mm, according to Gauss's theorem the electric field in air at the boundary of copper and air can easily exceed 1kV/mm. This high field causes local partial discharges.

During generation of ignition pulses the boost converter will operate in discontinuous conduction mode. The mosfets of the boost converter and the HV-circuit can operate with the same duty-cycle. With the previously mentioned duty-cycle of 25% and an assumed boost voltage of 50V results for the average input current and the battery power:

\[
I_{av} = \frac{I_{max}}{2} \left( \frac{D \cdot D}{V_c - V_i} \right) \quad I_{av} = \frac{V_{bat} \cdot DT}{L_{boost} \cdot 2} \left( \frac{D + 12}{38} \right) = 0.3A \quad P_{bat} = 3.6W
\]

The energies in the take-over capacitor and the ignitor capacitor (68nF) both charged with their nominal voltage are:
The pulse frequency with an estimated total efficiency of 50% and a supposed duty-cycle of 25% is:

\[ f_p = \frac{3.6W \cdot 0.5 \cdot \frac{1}{37.1mJ}}{\mu F \cdot (175V)^2} = 49Hz \]  

A higher pulse rate can be obtained easily through a higher duty-cycle. The maximum switching frequency of the spark gap is 400Hz, this results in a maximum battery input power of 30W. The ignitor voltage and take-over voltage are measured with a fixed duty-cycle of 25% for both the converters. The connection between the boost converter and the commutator is omitted for this set-up (see figure 3.1). The lamp only flashes with this repetitive ignition and take-over.

This spark gap breaks down at 770V, the OCV is 185V, the boost voltage is 45V and the \( f_p \) is 40. Performance is satisfying and comes close to calculation.
6 COMMUTATOR

The commutator transforms the DC voltage into a low frequency square wave voltage. The commutator contains a full bridge circuit with integrated circuits for gate driving and the current limitation for take-over (in case the impedance of the series ignitor is low), see figure 6.1:

![Diagram of the commutator](image)

Figure 6.1: Commutator

6.1 GATE DRIVING

Driving of the high mosfet in a half bridge is difficult because its source is floating in relation with the ground. Level shifting is required. Magnetic components or high voltage transistors are needed for level shifting in case the gate driving circuits are built with discrete components.

The availability of a high voltage IC process has led to a design of a HV-IC for gate driving. Philips has a HV-IC process in DMOS technology with lateral double diffused mosfets (LDMOS) for high voltage applications (Ludikhuize [34]). A gate driver IC for a half bridge application is designed by Schoofs [33]. This
The commutating frequency has to be set by an oscillator circuit. Our competitor International Rectifier makes a mos-gate driver (MGD) with an incorporated oscillator for a half bridge application [13]. This selected IR2155 MGD is widely available for an aggressive price. The IR2155 switches according to the break before make principle. The voltage commutation of the half bridge occurs in 1.2μs. A functional block diagram of the internal circuitry is depicted in figure 6.2.

\[ V_{\text{max}} = 600\text{V}, \quad I_{\text{source}} = I_{\text{sink}} = 500\text{mA}, \quad \text{Undervoltage Lockout} = 8\text{V}, \quad \text{DT} = 1.2\mu\text{s} \]

Deadtime: LS turn-off to HS turn-on or HS turn-off to LS turn-on.

![Figure 6.2: IR2155 mos-gate driver with incorporated oscillator](image)

The oscillation frequency is set by an external resistor (connected from RT to CT) and capacitor (connected from CT to
COM), the commutating frequency is:

\[ f_{osc} = \frac{1}{1.4 \times R_x \times C_T} = \frac{1}{1.4 \times 82 \, \Omega \times 33 \, \mu F} = 264 \text{Hz} \]  

(83)

The other bridge leg is driven by an IR2111, this is a half bridge driver similar to the IR2155 except without an oscillator. The turn-off propagation time is 150ns, the turn-on propagation time is 850ns (break before make). Voltage commutation of the half bridge leg driven by the IR2111 occurs in 0.7\mu s. The \( R_x \) terminal of the IR2155 is connected to the input of the IR2111 which results in a master slave connection.

To lock the bridge during ignition or for postponing the first current commutations (for obtaining a reliable ignition with 200V open circuit voltage) the capacitor \( C_T \) can be short-circuited with a small signal mosfet or bipolar transistor.

6.2 BRIDGE MOSFETS AND CAPACITORS

Ignition of the high-pressure MHS gas-discharge lamp occurs with an extremely high \( \frac{dI}{dt} \) since the mechanism for breakdown is a streamer. This high \( \frac{dI}{dt} \) causes a high \( \frac{d\Phi}{dt} \) field which induces undesired voltages inside the ballast and especially inside the commutator, see figure 6.3.

![Figure 6.3: Consequence of ignition](image)

In case there is no conducting path between both the outputs of the commutator (A and B) for letting flow a current induced by this \( \frac{d\Phi}{dt} \); the voltage in the commutator will rise to a high value and will certainly destroy the bridge mosfets.
The technical best solution is to make a good conducting circuit in which the induced current can flow and with a low transfer impedance to the existing commutator ignitor connection. For instance a connection with two coax cables. The cores of the coax cables are the two existing connections wires. The two sleeves are connected with each other at both ends of the cables. Carefully attention has to be paid to the construction at the cable ends for achieving a low transfer impedance. Unfortunately this is too cost consuming and is not suitable in an automotive environment.

The solution chosen is to reduce the induced voltage in the existing connection by adding capacitors in the commutator (depicted in figure 6.1) and using mosfets able to dissipate avalanche energy. The capacitors do have a double function, they create a reactive conducting path for an induced high frequency current. And the capacitors also reduce the dv/dt of the slope of the half bridge output during the voltage commutation for achieving a lower electromagnetic interference. The capacitors selected are two 10nF/250V MKT film capacitors.

The bridge mosfets used are manufactured by International Rectifier. IR manufactures a mosfet (a HEXFET process) which allows a relatively high single pulse avalanche energy ($E_{AS}$). The mosfet selected is a compromise between price and power loss (efficiency) caused by the $R_{ds\,on}$.

<table>
<thead>
<tr>
<th></th>
<th>$V_{ds,max}$ (V)</th>
<th>$I_{ds,max}$ (A)</th>
<th>$R_{ds,on}$ (Ω)</th>
<th>$E_{AS}$ (mJ)</th>
<th>50k</th>
<th>500k</th>
</tr>
</thead>
<tbody>
<tr>
<td>IRF620</td>
<td>200</td>
<td>5.2</td>
<td>0.80</td>
<td>110</td>
<td>$0.34$</td>
<td>$0.32$</td>
</tr>
<tr>
<td>IRF630</td>
<td>200</td>
<td>9</td>
<td>0.40</td>
<td>250</td>
<td>$0.46$</td>
<td>$0.43$</td>
</tr>
<tr>
<td>IRF640</td>
<td>200</td>
<td>18</td>
<td>0.18</td>
<td>580</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

The IRF630 is selected, the gate charge and capacitance graphs are given in figure 6.4.
Experiments were satisfying and showed that the induced energy caused by ignition is partly dissipated in the mosfets without destroying them. The power dissipated in the bridge mosfets caused by the \( R_{ds_{on}} \) is:

\[
P_{\text{bridge}} = 2 \cdot I_{\text{lamp}}^2 \cdot R_{ds_{on}} = 0.4 W
\]  

(84)

The efficiency of the ballast in steady-state operation for 12V input voltage and 50V lamp voltage can be calculated as follows:

\[
\eta_{\text{ballast}} = \eta_{\text{boost}} \frac{35 W - P_{\text{diode}} - P_{\text{bridge}}}{35 W} = 82\%
\]  

(85)

6.3 CURRENT LIMITATION FOR CONTROLLED TAKE-OVER

The discharge of the take-over capacitor can be controlled with a current limitation incorporated in the commutator. The current limitation circuit \([32,33]\) is depicted in figure 6.1. The resistor between the source of the bridge mosfet and the ground determines the maximum bridge current.

\[
I_{\text{max}} = I_{\text{take-over}} = \frac{V_{\text{be}}}{R_s} + I_{\text{b}} = \frac{0.7 V}{R_s}
\]  

(86)

A maximum bridge and take-over current of 10A is achieved with a 70m\( \Omega \) resistor. With this resistor the lamp current can be measured easily for the control of the ballast.
7 IGNITOR

The ignitor creates a conducting breakdown channel inside the lamp. The voltage required for a hot-restrike is 25kV with a fast risetime. The electric circuit of the ignitor with a spark gap as switching element is depicted in figure 7.1.

![Figure 7.1: Series ignitor with lamp](image)

7.1 CAPACITORS AND SPARK GAP

The capacitance choice of Cign is based on a minimum required ignition energy. Experiments showed that the minimum capacitance able to generate 25kV with a 800V spark gap is 68nF. The one chosen is a 68nF/1kV MMKP polypropylene film capacitor. The ESR and capacitance are measured as function of the frequency.

![Figure 7.2: Capacitance and ESR of the ignitor capacitor](image)
The spark gap is manufactured by Siemens and its breakdown voltage has a large tolerance of 15%. Unfortunately the spark gap is quite expensive.

A capacitor $C_d$ is placed parallel to the lamp and secondary winding of the ignitor. This capacitor has a low impedance for the high frequency ignition pulse. The capacitor decouples the ignition voltage from the commutator output voltage. The capacitor selected is a Murata 10nF/1kV ceramic capacitor.

7.2 CUSTOM-MADE TRANSFORMER

The step in voltage which has to be achieved is high, from 800V up to 25kV. A high turns ratio and a good couple factor are required. The turns ratio chosen is 1 to 40, based on experiments. This turns ratio would result for an ideal transformer in an ignition voltage of 32kV. To obtain a good coupling with this high turns ratio a foil is used for the primary winding. The transformer contains the following parts:
- ferrite rod as magnetic core
- custom-made coilformer
- inside primary winding, a foil
- outside secondary winding, a solid wire with 0.45mm diameter.

The number of primary turns is 3, the number of secondary turns is 120. The primary and secondary inductance and resistance are measured at 1kHz:

$L_1(\text{sec open})=0.91\mu\text{H}$ \quad $R_1=0.06\Omega$

$L_2(\text{prim open})=415\mu\text{H}$ \quad $R_2=0.95\Omega$

The secondary inductance and resistance are high. The resistance causes power loss and dissipation in the ignitor. The high secondary inductance is disadvantageous for a fast current commutation and fast take-over.
The capacitive energy in the take-over capacitor (equation 80) can cause a maximum take-over current of 8.6A:

\[ E_{so} = 15.3mJ = \frac{1}{2} \times 415\mu H \times (8.6A)^2 \]

(87)

The current limitation is superfluous with this series ignitor and is therefore omitted (figure 8.1). The ignitor in the experimental set-up is placed in oil, otherwise breakdown at the secondary side would occur. The definitive ignitor will be potted with an insulating material with a low relative dielectric constant \((\varepsilon_r)\). The generated ignition voltage and the ignitor capacitor voltage are measured at breakdown of the spark gap, see figure 7.3.

![Figure 7.3: Ignition voltage](image)

The upper graph is the ignition voltage with a maximum peak value of 25kV. The lower graph is the voltage over the ignitor capacitor. The frequency of the current flowing through the primary inductance of the transformer is determined by the values of the leakage and mutual inductances, the ignitor capacitor, the parasitic secondary capacitance and the impedance of the spark gap (breakdown of the spark gap is visible). The frequency measured is
909kHz. An ideal parallel connection of $L_1$ and $C_{ign}$ would result in a lower frequency of 640kHz. The spark gap does not turn-off at the zero crossing of the current.

The ignition voltage is an addition of the 909kHz voltage (main component) and a 3.3MHz voltage. This 3.3MHz voltage component is caused by the leakage inductance and the parasitic capacitance of the secondary winding.
8 SYSTEM PERFORMANCE AND RECOMMENDATIONS

A number of successful experiments has been done with old TCS lamps [5,10]. The differences in respect with the new lamp are an electrode diameter of 300μm, a 6bar xenon filling and an other metal halide division: So/Sc/Th=76/19/5. The mercury dose is the same: 0.1mg.

8.1 EXPERIMENTAL SET-UP

Due to the absence of a well matched feedback control the following set-up is invented. The ballast ignitor lamp configuration according to figure 8.1 is built and a resistor (extra load) is added to the output of the boost converter. The resistor is the load for the boost converter in case the lamp does not burn or the lamp suddenly extinguishes, the output voltage will not exceed the 60V limit with this resistor. The ballast is supplied with a power supply (the battery) with an adjustable output current limitation. The boost converter with resistor load is operating with a fixed duty-cycle in continuous conduction mode. The output voltage in this condition has to be higher than the steady-state lamp voltage, the current limitation of the power supply is not activated.

Ignition occurs through activating the HV-circuit. The activated HV-circuit operates with the same switching frequency and duty-cycle as the boost converter. After a successful ignition, take-over and run-up the HV-circuit is switched off (this means the gate of the mosfet is at ground level). The lamp current is limited by the adjustable current limitation of the power supply. The experimental ballast is powered by a current source.
Figure 8.1: Ballast ignitor lamp configuration
8.2 SYSTEM MEASUREMENT

The voltage of one output of the commutator and the lamp current are measured during ignition, see figure 8.2.

![Figure 8.2: Commutator output voltage and lamp current](image)

The upper graph is the voltage, the OCV is 170V at the moment that ignition occurs. Take-over current is flowing and the lamp voltage drops to about 30V. The current supplied by the boost converter is about 3A (lower graph), the run-up power equals 90W. The probability that the lamp extinguishes at the first current commutations is present. In case the lamp extinguishes the bridge voltage rises until the moment that the combination of electrode voltage and cathode temperature is high enough for emitting electrons and re-ignition occurs. Re-ignition with increased voltage (110V) is showed at the eighth current commutation in figure 8.2. Ignition, take-over and run-up are successful, the lamp voltage stabilises at 48V and the 35W lamp power is controlled with the output current limitation of the power supply, see figure 8.3. The commutation frequency is in accordance with equation (83).
8.3 RECOMMENDATIONS

8.3.1 Lower cost

The spark gap is the most expensive component of the ballast and ignitor. A gate turn-off thyristor as switching element could result in a lower cost solution. Use of a GTO creates the possibility of synchronizing ignition and current commutation.

The power mosfet in the boost converter can be replaced by a lower cost version with a higher $\text{R}_{\text{ds(on)}}$. For instance the: STP40N06: $V_{\text{ds(max)}}=60\text{V}$, $I_{\text{ds(max)}(25^\circ\text{C})}=40\text{A}$, $R_{\text{ds(on)}}=0.035\Omega$, $C_{\text{iss}}=1130\text{pF}$ or STP30N06: $V_{\text{ds(max)}}=60\text{V}$, $I_{\text{ds(max)}(25^\circ\text{C})}=30\text{A}$, $R_{\text{ds(on)}}=0.050\Omega$, $C_{\text{iss}}=950\text{pF}$. The lower input capacitance results in a faster switching speed. The lower maximum drain current and the higher $R_{\text{ds(on)}}$ will cause problems, especially at higher temperatures and during run-up.

The take-over capacitor is overdimensioned and can be replaced by a 200V capacitor with a smaller capacitance. The lowest capacitance possible will be determined by the control of the ballast.
8.3.2 Higher efficiency

The 3C80 core of the inductor of the boost converter can be replaced by a 3F3 core. This 3F3 ferrite has a lower specific power loss (figure 4.6). The core losses with 3F3 will be 0.13W (equation 43) instead of 0.39W.

The turn-off switching losses of the mosfet of the boost converter can be reduced by resonant switching. Resonant switching requires more components and therefore a higher ballast price.

8.3.3 Better reliability

The first current commutations after ignition are the most critical moments for the lamp to extinguish. To achieve a more reliable run-up with a relatively low open circuit voltage the first current commutations have to be delayed. With this postponed current commutation one achieves a current commutation with warmer electrodes (in particular the cathode). So the probability of extinguishing at the first current commutations will be reduced considerably.
9 PRICE CALCULATIONS

The prices of the components of the power part of the ballast and the ignitor are presented. All prices are component or material prices without costs for mounting or housing and without the printed circuit board. The prices per component are valid for a quantity in the order of 100k per year.

9.1 BOOST CONVERTER

The input capacitors are changed, the old X7R ceramic capacitors were far too expensive and are replaced by Z5U ceramic capacitors. The capacitance of Z5U material is more voltage dependent. The performance of the boost converter remains unchanged since the battery voltage does not change very much.

Inductor: core, 2*DFL 0.10
  coilformer
  Litz-wire
Mosfet: SGS-Thomson, STP50N06  DM 1.20 → DFL 1.36
Diode: Philips, BYW29-100
Boost capacitor: Philips, 220nF/63V MKT
Input capacitor: Siemens, 1μF/63V Z5U, 3*DFL 0.17

The price of the boost converter is DFL 3.16.

9.2 TAPPED-INDUCTOR BOOST CONVERTER

Transformer: core, 2*DFL 0.15
  coilformer
  solid wire
Mosfet: International Rectifier, IRF610  $ 0.30 → DFL 0.58
Diodes: Philips, BYV26 D(800V)
  Philips, BYV26 B(400V)
Take-over capacitor: Siemens, 1μF/250V MKP
Connections diodes: Philips, BYV27-200  DFL 0.28
   Philips, BYV27-100  DFL 0.25

The price of the tapped-inductor boost converter together with the connection diodes is DFL 2.81.

9.3 COMMUTATOR

Gate driver ICs: International Rectifier, IR2155 $ 0.80 → DFL 1.55
                   International Rectifier, IR2111 $ 0.79 → DFL 1.53
Mosfets: International Rectifier, IRF630 4*$0.46 $ 1.84 → DFL 3.56
Capacitors: Philips, 10nF/250V MKT, 2*DFL 0.10       DFL 0.20

The price of the commutator is DFL 6.84.

In the future the IR2155 has to be replaced by the then available IR2151. Price of the IR2151 is $ 0.55. The technical characteristics of the IR2151 are almost similar to those of the IR2155 and the performance of the commutator does not change with this lower cost version.

9.4 IGNITOR

Transformer: core     DFL 0.40
                coilformer, custom-made  DFL 0.50
                foil                   DFL 0.35
                solid wire           DFL 0.20
Spark gap: Siemens, 800V       DM 3.00 → DFL 3.41
Ignitor capacitor: Philips, 68nF/1kV MMKP  DFL 0.53
Decouple capacitor: Murata, 10nF/1kV           DFL 0.18

The price of the ignitor is DFL 5.57.
9.5 PRICE COMPARISON WITH PRESENT MPL SYSTEM

The prices of the power parts of the low cost ballast and the MPL ballast are considered and compared. All prices in this comparison are component or material prices without costs for mounting or housing and without the printed circuit boards. The prices of the MPL ballast are calculated in the same way and for the same quantity as the prices of the low cost ballast.

<table>
<thead>
<tr>
<th>Component</th>
<th>Low cost ballast</th>
<th>MPL ballast</th>
</tr>
</thead>
<tbody>
<tr>
<td>Converter:</td>
<td>DFL 3.16</td>
<td>DFL 22.60</td>
</tr>
<tr>
<td>HV-circuit:</td>
<td>DFL 2.81</td>
<td>-</td>
</tr>
<tr>
<td>Re-ignition coil:</td>
<td>-</td>
<td>DFL 2.45</td>
</tr>
<tr>
<td>Commutator:</td>
<td>DFL 6.84</td>
<td>DFL 15.80</td>
</tr>
<tr>
<td>Ignitor:</td>
<td>DFL 5.57</td>
<td>DFL 6.71</td>
</tr>
<tr>
<td><strong>Total:</strong></td>
<td><strong>DFL 18.38</strong></td>
<td><strong>DFL 47.56</strong></td>
</tr>
</tbody>
</table>

This means a total price reduction of the components of the power part with a factor 2.6.

A lot of profit is achieved with the introduction of the boost converter. This makes clear the necessity for the 50V lamp. The HV-circuit is at the same price level as the re-ignition coil. A commutator with integrated circuits for gate driving appears to be a low cost solution. The 800V ignitor works out less expensive than the 1600V ignitor due to the lower cost of the custom-made transformer.
10 CONCLUSIONS

The presented low cost ballast powers a MHS gas-discharge lamp with 50V lamp voltage. This 50V lamp obtained by a reduced mercury dose is absolutely necessary for achieving a ballast with a low price and a small volume.

The performed development strategy has resulted in a new ballast approach: a main high efficiency converter with limited output voltage for the power conversion and a second converter for generating the ignition voltage and the open circuit voltage.

A good ballast performance without a re-ignition coil is obtained with a power converter with a fast transient response and a controlled capacitor discharge for take-over.

The component price of the power part of the ballast is DFL 12.81, the component price of the ignitor is DFL 5.57. Both prices are without costs for mounting and housing.

The ballast is designed with two small magnetic components: an U11 inductor and an E13 transformer.

The efficiency of the ballast is 82% for a 12V battery voltage and 35W lamp power. The efficiency increases if the ratio of lamp voltage and battery voltage decreases.
## 11 DEFINITIONS OF USED SYMBOLS

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Definition</th>
</tr>
</thead>
<tbody>
<tr>
<td>V, v</td>
<td>Voltage (V)</td>
</tr>
<tr>
<td>( V_{\text{nom}} )</td>
<td>Nominal voltage (V)</td>
</tr>
<tr>
<td>( V_{\text{bat}} )</td>
<td>Battery voltage (V)</td>
</tr>
<tr>
<td>Vin, Vi</td>
<td>Input voltage (V)</td>
</tr>
<tr>
<td>Vout, Vo</td>
<td>Output voltage (V)</td>
</tr>
<tr>
<td>I, i</td>
<td>Current (A)</td>
</tr>
<tr>
<td>( I_{\text{av}} )</td>
<td>Average current (A)</td>
</tr>
<tr>
<td>( I_{\text{rms}} )</td>
<td>Effective current, root mean square current (A)</td>
</tr>
<tr>
<td>I, I</td>
<td>Inductor current (A)</td>
</tr>
<tr>
<td>Iin, Iout</td>
<td>Input current (A)</td>
</tr>
<tr>
<td></td>
<td>Output current (A)</td>
</tr>
<tr>
<td>P, P</td>
<td>Power (W)</td>
</tr>
<tr>
<td>Pin, Pout</td>
<td>Input power (W)</td>
</tr>
<tr>
<td></td>
<td>Output power (W)</td>
</tr>
<tr>
<td>T, t</td>
<td>Time (s)</td>
</tr>
<tr>
<td>f, w</td>
<td>Frequency (Hz)</td>
</tr>
<tr>
<td>s</td>
<td>Angular frequency (1/s)</td>
</tr>
<tr>
<td>Q</td>
<td>Laplace's quantity (1/s)</td>
</tr>
<tr>
<td>E</td>
<td>Charge (C)</td>
</tr>
<tr>
<td>E</td>
<td>Energy (J)</td>
</tr>
<tr>
<td>D, d</td>
<td>Duty-cycle</td>
</tr>
<tr>
<td>M</td>
<td>Voltage gain</td>
</tr>
<tr>
<td>n</td>
<td>Efficiency</td>
</tr>
<tr>
<td>C</td>
<td>Capacitance (F)</td>
</tr>
<tr>
<td>ESR</td>
<td>Equivalent series resistance (Ω)</td>
</tr>
<tr>
<td>R, r</td>
<td>Resistance (Ω)</td>
</tr>
<tr>
<td>Rl, R</td>
<td>Inductor resistance (Ω)</td>
</tr>
<tr>
<td>Rc, Rm</td>
<td>Capacitor resistance (Ω)</td>
</tr>
<tr>
<td>Rd</td>
<td>Mosfet resistance (Ω)</td>
</tr>
<tr>
<td>L, Le</td>
<td>Diode resistance (Ω)</td>
</tr>
<tr>
<td></td>
<td>Inductance (H)</td>
</tr>
<tr>
<td></td>
<td>Effective inductance (H)</td>
</tr>
<tr>
<td>Symbol</td>
<td>Description</td>
</tr>
<tr>
<td>--------</td>
<td>-------------</td>
</tr>
<tr>
<td>L1</td>
<td>Primary inductance (H)</td>
</tr>
<tr>
<td>L2</td>
<td>Secondary inductance (H)</td>
</tr>
<tr>
<td>Lm</td>
<td>Mutual inductance (H)</td>
</tr>
<tr>
<td>k</td>
<td>Coefficient of coupling</td>
</tr>
<tr>
<td>J</td>
<td>Current density (A/m²)</td>
</tr>
<tr>
<td>E</td>
<td>Electric field intensity (V/m)</td>
</tr>
<tr>
<td>D</td>
<td>Electric flux density (C/m²)</td>
</tr>
<tr>
<td>H</td>
<td>Magnetic field intensity (A/m)</td>
</tr>
<tr>
<td>B,b</td>
<td>Magnetic flux density (Vs/m²)</td>
</tr>
<tr>
<td>Φ</td>
<td>Magnetic flux (Vs)</td>
</tr>
<tr>
<td>μ</td>
<td>Permeability (Vs/Am)</td>
</tr>
<tr>
<td>μ₀</td>
<td>Permeability of vacuum (4.π.10⁻⁷ Vs/Am)</td>
</tr>
<tr>
<td>μᵣ</td>
<td>Relative permeability</td>
</tr>
<tr>
<td>Ae</td>
<td>Effective area (m²)</td>
</tr>
<tr>
<td>lₐ</td>
<td>Air gap (m)</td>
</tr>
<tr>
<td>N</td>
<td>Number of turns</td>
</tr>
<tr>
<td>Pfe</td>
<td>Power losses in ferrite (W)</td>
</tr>
<tr>
<td>Pcμ</td>
<td>Power losses in copper (W)</td>
</tr>
<tr>
<td>Vds</td>
<td>Drain source voltage (V)</td>
</tr>
<tr>
<td>Ids</td>
<td>Drain source current (A)</td>
</tr>
<tr>
<td>Rds_on</td>
<td>Drain source resistance of a conducting mosfet in the ohmic region (Ω)</td>
</tr>
<tr>
<td>Qg</td>
<td>Gate charge (C)</td>
</tr>
<tr>
<td>Ciss</td>
<td>Input capacitance (F)</td>
</tr>
<tr>
<td>Cds</td>
<td>Drain source capacitance (F)</td>
</tr>
<tr>
<td>Eas</td>
<td>Single pulse avalanche energy (J)</td>
</tr>
<tr>
<td>Vₓ</td>
<td>Forward voltage (V)</td>
</tr>
<tr>
<td>Vₓ</td>
<td>Reverse voltage (V)</td>
</tr>
<tr>
<td>Iₓ</td>
<td>Forward current (A)</td>
</tr>
<tr>
<td>Qs</td>
<td>Stored charge (C)</td>
</tr>
<tr>
<td>tₓₓ</td>
<td>Reverse recovery time (s)</td>
</tr>
<tr>
<td>Vbe</td>
<td>Basis emitter voltage (V)</td>
</tr>
</tbody>
</table>
## 12 Definitions of Used Abbreviations

<table>
<thead>
<tr>
<th>Abbreviation</th>
<th>Definition</th>
</tr>
</thead>
<tbody>
<tr>
<td>AC</td>
<td>Alternating current</td>
</tr>
<tr>
<td>CCM</td>
<td>Continuous conduction mode</td>
</tr>
<tr>
<td>DC</td>
<td>Direct current</td>
</tr>
<tr>
<td>DCM</td>
<td>Discontinuous conduction mode</td>
</tr>
<tr>
<td>DFL</td>
<td>Dutch florin</td>
</tr>
<tr>
<td>EMI</td>
<td>Electromagnetic interference</td>
</tr>
<tr>
<td>HV</td>
<td>High voltage</td>
</tr>
<tr>
<td>MGD</td>
<td>Mos gate driver</td>
</tr>
<tr>
<td>MHS</td>
<td>Motorcar headlight system</td>
</tr>
<tr>
<td>Mosfet</td>
<td>Metal oxide semiconductor field effect transistor</td>
</tr>
<tr>
<td>MPL</td>
<td>Micro power light</td>
</tr>
<tr>
<td>OCV</td>
<td>Open circuit voltage</td>
</tr>
<tr>
<td>PWM</td>
<td>Pulse width modulation</td>
</tr>
<tr>
<td>TCS</td>
<td>Totally combined system</td>
</tr>
<tr>
<td>Vedilis</td>
<td>Vehicle discharge light system</td>
</tr>
</tbody>
</table>
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