Relación de transformación óptima para fuente de alimentación de lámpara de descarga de barrera dieléctrica

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Optimum Transformer Turns Ratio of Power Supply for Dielectric Barrier Discharge Lamp

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1

Introduction

Investigations of the bactericidal effect of sunlight in the 19th century planted the seed of disinfection by ultraviolet (UV) radiation [1]. Since then, UV attains growing application as an antimicrobial agent. Water disinfection is currently the most advanced and accepted germicidal application [2] although also it is used to disinfect air, surfaces. Due to the photochemical reactions generated by the UV radiation, it is even applied in industrial and medical processes.

The main advantage of UV sterilization and disinfection methods, as compared to those using chemical agents (chlorine, hydrogen peroxide, ozone, etc.), consists in fact that the radiation does not produce any toxic residuals [2].

There are several technologies to artificially produce UV, and according to the application a solution based on a type of UV source can be designed. For germicidal systems, High Intensity Discharge Mercury based lamps are widely used. However, due to the environmentally unfriendly nature of the mercury, the research on different UV sources as Dielectric Barrier Discharge (DBD) excimer lamps has gained importance.

In order to generate a discharge, DBD consists in applying a high enough voltage to a static gas where the electrodes are separated from the gas by at least one dielectric barrier. Because of the dielectric barrier, the electric arc is avoided and a flux of ions is generated.

As the electrical behaviour of this type of lamps has a capacitive nature, for a correct performance, the lamp requires a power supply with a symmetrical bipolar current waveform to prevent an uncontrolled voltage across the lamp [3].

Several topologies are known to drive DBD loads in current mode. In the research preceding this project [4] different power supplies were designed and compared, all converters of [4] as well as most of the topologies found in the literature, uses a step-up transformer which easily provides the high ignition voltage (several kV).

[5], [6], [7] and [3] reported that by connecting the transformer directly to the lamp, the transformer parasitic effects alter the system behaviour. One of the most relevant effects is produced by the transformer equivalent capacitor, this element causes that an important among of the output current flows through the transformer instead of the lamp. This may even affect the ignition of the lamp.

Based on the presented difficulties by converters with transformers, some works have been done about the design of high voltage transformers to reduce the parasitic capacitive effects [8]. Although the improvements in the windings arrangement lead to a better operation of the power supplies, [4] hypothesizes that the development of a transformer-less DBD power supply could reduce the supply size while increasing its efficiency and reliability.
1. Introduction

The principal objective of this project is: to design and implement a power supply for a DBD excimer lamp with the optimum transformer turns ratio, considering the transformer-less case. In order to reach the main objective, the following specific objectives were set: (1) to study different topologies for driving DBD lamps and select one; (2) to implement the chosen topology; (3) to evaluate the results throughout an experimental or theoretical comparison with other power supplies.

Moreover, as there are important precedents on the effects of the parasitic components, it is necessary to analyze the selected topology under the influence of the parasitic capacitances.

In order to provide a complete description of the relevant aspects associated with the development of this project, the following subjects are described as follow:

In the first chapter the theoretical framework is presented. It contains general concepts regarding the DBD excimer lamps and its power supplies.

The second chapter presents the analysis, design and implementation of the selected topology without transformer. First, the analysis of the circuit will be explained in detail, then the high voltage switch is presented, following by the hardware implementation and results.

As the results of the transformer-less power supply show an extremely low efficiency caused by the parasitic capacitance, in the third chapter the effects of the switches capacitances are analysed. Additionally, a relation between the losses and the transformer turns ratio is stabilised in order to find an optimum converter and compare it with the one without transformer.

Based on the objectives of this work and its results, in the final chapter the conclusions are presented.
2

Theoretical Framework

2.1 Overview of DBD Excimer Lamps

DBD excimer lamps exhibit various advantages over conventional UV sources. As these lamps are mercury-free, they are a solution for moving away from mercury UV sources[9]. Severe mercury exposure can lead to damage to human brain cells, the kidneys, the liver and the nervous system, that is why mercury products are becoming more and more restricted around the world.

Although the environmental aspects of the presence of mercury is an important concern, the lighting industry still use it since mercury products are, in most cases, more efficient than mercury-free products [9].

Currently, different options for light sources without mercury are being investigated and DBD excimer lamps provide a compelling solution that is strongly researched.

Besides the absence of mercury, DBD excimer lamps do not require a preheating time, and offer a longer lifetime given that the electrodes are not in direct contact with the gas, so they are wear-free.[10]

Additionally, DBD excimer lamps present a very spectral-selective UV production which depends mainly on the gas composition[11]. This allows a greater selectivity of the radiated wavelength, for example, excimer molecules of $XeCl^*$ produce UV at 308 nm, and $KrCl^*$ at 222 nm.

![Figure 2.1: DBD excimer lamp](image)
2. Theoretical Framework

2.1.1 Operating Principle

The working principle of DBD Excimer lamps is based on the generation of excimer molecules by means of Dielectric Barrier Discharges in which by applying a high voltage between the electrodes of the lamp, a flux of ions is generated[12]. The DBDs are characterized by the presence of a dielectric layer between the electrodes and the gas. The purpose of the dielectric walls is to limit the current between the electrodes[9]. When a high enough voltage establishes an avalanche of electrons, occurs the transition to a self-sustained current, which flows through the ionized channel. This conducting are called streamers.

DBDs can produce an ideal environment for excimer generation. Excimers are transient species that are weakly bound in excited states of molecules[12]. If the gas is composed only by one element, the created specie is called an excimer, and if the gas is composed by two elements, the excited molecule is called an exciplex.

Excimers have attractive features for UV production. Due to the transition between a weakly bound excited state to a ground state, they emit light in the form of ultraviolet radiation.

The UV production in DBD excimer lamps can be summarized as follow: when an avalanche of electrons produced by an electric field between the lamp electrodes is initiated, the electrons collide with the gas atoms or molecules in such a way that the molecules are associated in an excited state of energy. Once the new excited molecule (excimer) is dissociated, it returns to its ground state and releases its energy as an emission of UV photons. Figure 2.2 shows the formation of an XeCl exciplex [12].

![Formation of an XeCl exciplex in a DBD excimer lamp](image)

**Figure 2.2:** Formation of an XeCl exciplex in a DBD excimer lamp [12]

2.1.2 DBD geometry and configuration

There are different configurations of DBD light sources. The planar setup is made by two parallel plates with at least one dielectric layer. For most of UV emission devices the cylindrical geometry is used, Figure 2.3 shows the structure of this configuration.

![Cylindrical DBD structure](image)

**Figure 2.3:** Cylindrical DBD structure a)longitudinal view b)cross-sectional view
2. Theoretical Framework

In the cylindrical geometry the gas is confined between two dielectric barriers, typically the dielectric materials used are glass, quartz, or ceramics[10]. In order to allow the flux of radiation out of the lamp, the external electrode is a metallic mesh that partially obstruct the UV radiation.

2.1.3 Electrical Modeling DBD Excimer Lamps

In order to design the power supplies and execute computer simulations, lamps can be replaced by their models. The most complete models of DBD lamps are focused on physical performance, they can predict the electrical behavior, as well as the spectral distribution of light, and even a description of spatial and time evolution of the discharge plasma [13] [14]. Since these models are very complex and are mainly used for the lamps design, they will not be considered in this project.

On the other hand, there are simplified models that are complete enough to analyse and design power supplies. They just describe the electrical behavior without chemical reactions in the plasma or energy fluxes.

As can be seen in Figure 2.4, the electrical characteristics of DBDs can be modeled by a simple circuit [15]. The capacitances of the external and internal dielectric walls are represented in a series-equivalent capacitor called $C_d$. The dielectric barriers are in series with the gas, that behaves itself as a capacitor $C_g$ (before the breakdown) in parallel with the gas conductance ($G$).

Figure 2.5 presents the static characteristic of the discharge. In steady state before the breakdown begins ($v_{gas} < V_{th}$), the gas presents a very low conductance, therefore the equivalent model will be the series connection between $C_d$ and $C_g$. While after the breakdown ($v_{gas} > V_{th}$), the gas leads to a constant voltage source of $V_{th}$ value [3].

The Lissajous figure method, reported by Manley is still a common method to experimentally determine the electrical DBD lamp parameters. Throughout this procedure, the charge/voltage Lissajous figure shows important information about the discharge. The Lissajous is close to and ideal parallelogram, from which the discharge voltage $V_{th}$ and the capacitances $C_g$ and $C_d$ can be inferred. Figure 2.6
presents an example of the charge/voltage plot of a DBD lamp and the relation between the slopes of the parallelogram with the model capacitances.

![Figure 2.6: Example of Charge-Voltage Lissajous figure [10]](image)

For the XeCl excimer lamp that intended to be used, the value of the capacitances $C_g$ and $C_d$ are in the order of tens of picofarads, 25 pF and 85 pF respectively, and the breakdown voltage is 1300 V [4].

### 2.2 Transformer-less Power Supplies

So far, a brief explanation about the operating principle and electrical model of the DBD excimer lamps have been presented. At this point, it is worth highlighting that DBD lamps are characterized by a capacitive nature, which requires a high voltage converter with symmetrical bipolar waveform to reach a discharge regime.

The first approximations in power supplies for driving DBD lamps used voltage sources. These converters only inject current to the lamp in case of a variation in the lamp voltage, consequently, it is difficult to predict the amplitude and duration of the lamp current waveform [16].

Otherwise with current sources, knowing the imposed lamp current it is possible to determine the lamp voltage by means of the equation 2.1.

$$v_{lp} = \frac{1}{C_d} \int i_{lp} dt + v_{gas}, \quad (2.1)$$

There are several topologies to drive DBD lamps in current-mode, most of them require a transformer to step-up the voltage in order to achieve the ignition.

As important issues with parasitic elements of the step-up transformer has been reported [5] [6], some converters will be analysed to determine the viability of their implementation without transformer.

#### 2.2.1 Converters review

In [16], it is stated a review of 4 different topologies to drive DBD excimer lamps in current mode. Following it is presented an analysis and comparison of these topologies in order to select one for the implementation without transformer.
2. Theoretical Framework

**Rectangular Current Converter** This converter is composed by a constant current source connected in cascade with a full bridge current inverter which load is the lamp. In [17] there is a detailed explanation of the operation of each block.

With the rectangular current converter, the duty cycle and frequency can be controlled in the current inverter, while the amplitude of the current injected into the lamp is controlled by the constant current source. The flexibility of this converter allows controlling the lamp power with three degrees of freedom, that is why it is used for the characterisation and identification of the optimal operating point of the lamp. Due to the eight switches in the converter, it is not optimum in terms of efficiency or cost [16] so it will not be considered.

**Boost-Based Converter** In this converter, the inductance L is charged linearly during a charging time $t_{ch}$, then there is the discharge subinterval, in which a resonance behaviour between the inductance and the equivalent capacitance of the lamp is established. Once the voltage on the gas reaches $V_{th}$, the resonance changes, since the equivalent capacitance of the lamp varies. According to the model of the Figure 2.4, the equivalent capacitance varies between $\frac{C_dC_g}{C_d+C_g}$ and $C_d$.

Only when the condition $V_{in} < V_{th}$ is fulfilled the converter will properly operate, otherwise, the energy in the lamp will grow uncontrollably [18].

This converter allow us to obtain a determined operating condition from a DC input voltage lower than the voltage required for the other evaluated converters [4].

![Figure 2.7: Boost Based Converter Topology](image1)

![Figure 2.8: Boost Based Current Waveforms](image2)

**Series Resonant Inverter** The SRI can be considered as an special case of the Boost-based converter where the charging time $t_{ch}$ is zero. As well as the Boost converter, it presents ZCS and in order to guarantee an stable operation, the input voltage must be lower than $V_{th}$. According to the results presented in [4], in comparison with the other three converters, this is the most efficient topology.
2. Theoretical Framework

**Buck–Boost Based Converter** The sequences of the Buck-Boost Converter are similar to the Boost ones. In this converter, the energy stored in the inductance, during the charge phase, is sent to the lamp during the discharge phase [16]. Unlike the SRI and Boost based converters, the buck-boost is stable for all the values of $V_{in}$.

**Table 2.1: Resonant-Based Converters Comparison**

<table>
<thead>
<tr>
<th>INVERTER</th>
<th>ADVANTAGES</th>
<th>DISADVANTAGES</th>
</tr>
</thead>
<tbody>
<tr>
<td>Boost-Based Converter</td>
<td>* Lower input voltage required</td>
<td>* Stability condition</td>
</tr>
<tr>
<td></td>
<td>* Four switches</td>
<td></td>
</tr>
<tr>
<td>Series Resonant Inverter</td>
<td>* Highest efficiency</td>
<td>* Stability condition</td>
</tr>
<tr>
<td></td>
<td>* Four switches</td>
<td></td>
</tr>
<tr>
<td>Buck–Boost-Based Converter</td>
<td>* Stable for all values of $V_{in}$</td>
<td>* Five switches - more expensive</td>
</tr>
<tr>
<td></td>
<td>* The power injected into the lamp is independent of the lamp parameters.</td>
<td></td>
</tr>
</tbody>
</table>
Table 2.1 presents a summary of the advantages and disadvantages of the resonant inverters previously explained to easily compare them.

When the transformer is removed from the converter, the input voltage is increased, that said, the Boost based resonant converter was chosen to design the power supply for DBD excimer lamps without transformer since it requires lower input voltages that can be achieved with laboratory equipment; the next chapter contains a thorough explanation of its design and working principle.

2.2.2 High Voltage Switches

As a consequence of removing the transformer, the stress voltages in the switches will be increased, therefore it will be necessary to use a suitable high voltage switch solution.

Silicon Carbide (SiC) power devices provide a compelling solution due to their high breakdown voltage, high operating temperature, high switching frequency and low losses.

At present, wide band-gap semiconductors are gaining more attention in high frequency and high voltage applications. Unfortunately, most of the high voltage devices are not commercially available and currently only high-performance 1.2kV SiC JFETs and SiC MOSFETs are possible to buy. Therefore, in order to extend the blocking voltage of a single SiC device, it is necessary to connect several devices in series to distribute the external voltage across this serial connection in the blocking state. The series stack of the devices can be implemented using SiC MOSFETs or JFETs.

Series Connection of MOSFETs  The series connection of MOSFETs require an adequate gate driver circuit to switch all the stacked MOSFETs with high timing precision. Additionally, a balancing network is used to equally distribute the voltage between the MOSFETs and limit the drain-source voltage during the transients and switching operation.

[19] uses the serial connection of MOSFETs for the design of a pulsed voltage source of 10 kV for Dielectric Barrier Discharges, Figure 2.13 presents its realized balancing network.

![Figure 2.13: Series stack of MOSFETs with balancing network][19]
2. Theoretical Framework

In the balancing network, the capacitances C avoid transient overvoltages and the resistances R guarantee static balancing. The zener diodes DZ are an additional overvoltage protection. This balancing network is a RCD snubber [19] [20].

The implementation of the driver is a challenge since the different delay times of the parts could lead to a none synchronous switching. The conventional driver with insulated DC-DC converters, optocouplers and IC-drivers is a clear example of this problem. Other methods as the transformer insulated gate driver, reduces the non-synchronous switching but has problems working at high duty cycles, as the transformer can deliver only ac signals, and the core flux must be reset each half cycle to maintain a volt-second balance [21].

**Series Connection of JFETs** Figure 2.14 shows the diagram of the series connection of JFETs, as can be seen in the figure, the chain of JFETs are connected in series with a low voltage MOSFET in a Super Cascode configuration.

![Series stack of JFETs](image)

**Figure 2.14:** Series stack of JFETs [22]

The Super Cascode is controlled via the gate of the MOSFET, so it can be easily driven. This configuration has a non-synchronous switching behavior which relates to a domino effect.

The static blocking characteristics is determined by the avalanche breakdown voltage of the diodes (D1 and D2), and the dynamic switching behaviour is improved using auxiliary resistors and capacitors for dynamically balancing the voltage distribution of the JFETs [23] [24].

Due to the relatively simple implementation of the driver in the Super Cascode JFET, it will be used for the design of the high voltage switches of the transformerless power supply. In section 3.2, it is presented a complete description of the operating principle and additional required considerations.
3

Transformer-less Boost Resonant Converter

3.1 Analysis and Design of Boost Based Resonant Converter

After reviewing the different topologies to drive DBD lamps, the boost based resonant converter introduced in [18], was selected for the implementation of a transformer-less power supply (Figure 3.1). This topology allows achieving the requested output power without transformer with a low input voltage.

Figures 3.3 - 3.10 show the different subintervals of commutation which are described in each semi-cycle as follows. The respective output waveforms can be seen in Figure 3.2.

(A) Charging Time (Figure 3.3 and Figure 3.7): During the time $t_{ch}$, the current $i_L$ increases linearly up to $I_{L0}$. The final charging current, is the initial current for the discharge sequence.

$$I_{L0} = \frac{V_{in}}{L} \times t_{ch}$$

(3.1)
(B) Discharge Subinterval - Before Ignition (Figure 3.4 and Figure 3.8): After the charging time, the lamp remains off. In this stage a resonant behaviour between $L$ and the equivalent capacitance of the lamp is established. As the gas has not reached yet the Threshold Voltage ($V_{TH}$), it presents a very low conductance.

$$\omega = \frac{1}{\sqrt{LC_{eq}}},$$

where $C_{eq} = \frac{C_d C_g}{C_d + C_g}$.

(C) Discharge Subinterval - After Ignition (Figure 3.5 and Figure 3.9): Once the voltage on the gas reaches $V_{TH}$, the gas behavior can be simplified as a voltage source of $V_{TH}$ value when the lamp current is positive, and $-V_{TH}$ if it is negative. In this subinterval the resonance changes, since the equivalent capacitance of the lamp is given only by $C_d$. The lamp will remain on until the change in the current direction.

$$\omega^* = \frac{1}{\sqrt{LC_d}}.$$

(D) All switches off (Figure 3.6 and Figure 3.10)

The negative cycle of operation (Figures 3.7, 3.8, 3.9 and 3.10) has the same sequence than the positive but reversing the direction of current to ensure the zero average current required by the lamp.
3.1.1 State plane analysis

The state plane is used to reduce the complexity of the resonant converters analysis. By properly normalizing the resonant tank waveforms, the problem can be described by segments of circles, lines, or other simple figures plotted in clockwise trajectories. This method can be used to derive the solutions for the steady-state output characteristics, to determine operating mode boundaries, and to find peak component stresses.

The waveforms are normalized using a base impedance equal to the tank characteristic impedance $R_0 = \sqrt{\frac{L}{C_{eq}}}$. The normalizing base voltage will be $V_N$. In the notation, the symbol for the normalized voltage is $X$, and $Y$ for the current.

$$X = \frac{v}{V_N}$$  \hspace{1cm} (3.2)  \\
$$Y = \frac{i}{V_N R_0}$$  \hspace{1cm} (3.3)  

It should be taken into account that after the ignition of the lamp (Subinterval C), the equivalent capacitance changes from $C_{eq}$ to $C_d$, so the normalization must be redefined with the new base impedance $R_0^* = \sqrt{\frac{L}{C_d}}$.

There are two cases to discuss in the state plane analysis. The first operating case occurs when the peak current is not reached before the lamp breakdown (Case A), while in the other (Case B), the ignition occurs once the current has already reached its maximum. According to the case, the behavior of the converter changes.

If the converter operates in equilibrium, then the trajectory begins and ends at the same point in the state plane, and the tank waveforms are periodic. Otherwise, a transient occurs in which the trajectory for each switching period begins at a different point, and follows a different path in the state plane. If the circuit is stable, then the trajectory eventually converges to a single closed path. To find the converter steady-state characteristics, we need to solve the geometry of this closed path. The steady state plane for each case is presented below.

**State plane in positive Subinterval B - Before Ignition**  In this stage, the applied tank voltage is equal to $V_{DC}$. Therefore the state plane trajectory is centered at $X_{DC}$, whose radius $r_1$ depends on the initial conditions.

As can be seen in Figure 3.2, at the beginning of the positive semicycle and after the time $t_{ch}$, the lamp voltage is $-V_{lp}$, and $I_{L0}$ is the initial current for the discharge sequence. Normalizing the initial condition we obtain that:

$$u(t_{ch}) = -V_{lp} = X_0$$  \hspace{1cm} (3.4)  \\
$$i_{lp}(t_{ch}) = I_{L0} = Y_{L0}$$  \hspace{1cm} (3.5)  \\
$$r_1^2 = (X_{DC} + X_0)^2 + Y_{L0}^2$$  \hspace{1cm} (3.6)
When the gas voltage is completely inverted from $-V_{TH}$ at $t_{ch}$ to $V_{TH}$ at $t_{br}$ the gas breakdown occurs. It means that the total gas voltage change in this subinterval is $2V_{TH}$.

\[ 2V_{TH} = \frac{1}{C_g} \int_{t_{ch}}^{t_{br}} i_{lp} dt \]  
(3.7)

\[ v_{lp}(t_{br}) = V_{lp}(t_{ch}) + \frac{1}{C_{eq}} \int_{t_{ch}}^{t_{br}} i_{lp} dt \]  
(3.8)

Replacing 3.7 in 3.8, and normalizing the expression, can be found the normalized lamp voltage at the breakdown time $t_{br}$

\[ X_R = -X_0 + 2X_{TH} \frac{C_g}{C_{eq}} \]  
(3.9)

From trigonometric equations, it is possible to obtain the current at the gas breakdown instant.

\[ Y_{R}^2 = (X_{DC} + X_0)^2 - (X_R + X_{DC})^2 \]  
(3.10)

**State plane in positive Subinterval C - After Ignition** The initial conditions of this stage, are equal to the current $Y_R$ and voltage $X_R$ from the previous subinterval. However, the normalization must be redefined since the equivalent capacitance changes. The equation to relate the currents of both state planes is:

\[ Y_{R}'^2 = Y_{R}^2 \frac{C_{eq}}{C_d} \]  
(3.11)

While the lamp is on, the gas voltage is $v_{C_g} = V_{TH}$, and from the lamp model we can obtain a relation between $X_{eq}$ and $X_d$:

\[ X_d = X_{eq} - X_{TH} \]  
(3.12)

The state planes trajectories for this subinterval are given in Figures 3.13 and 3.14, this subinterval begins at $\omega_0 t = t_{br}$, and ends when the current reaches zero.
The initial values of tank inductance current and capacitor voltage, are \( Y_R^* \) and \( X_R - X_{TH} \) respectively. And the center of the trajectory is \( X_{DC} - X_{TH} \).

From geometrical relations we can obtain that

\[
r_2 = X_0 - X_{DC}
\]  

\[(3.13)\]

**Figure 3.13:** State plane after ignition operating in Case A.  
**Figure 3.14:** State plane after ignition operating in Case B.

In order to determine the peak voltage expression, the currents at the gas breakdown \( Y_R \) and \( Y_R^* \) are equalized using the relations 3.11 and 3.9.

\[Y^2 = Y_{L0}^2 + (X_{DC} + X_0)^2 - (X_{DC} - X_R)^2\]  

\[(3.14)\]

\[Y^*^2 = (X_0 - X_{DC})^2 - (X_{DC} - X_R)^2\]  

\[(3.15)\]

Equating 3.14 and 3.15,

\[X_0 = \frac{C_d C_g Y_{L0}^2 - 4X_{TH}(C_g + C_d)(C_d (X_{DC} - X_{TH} - C_g X_{TH}))}{4C_d(C_g + C_d)(X_{TH} - X_{DC})}\]  

\[(3.16)\]

Denormalizing the equation 3.17:

\[\tilde{V}_{lp} = \frac{I_{L0}^2 L - 4V_{th}(V_{dc} C_g - V_{th}(C_g + C_d))}{4C_d(V_{th} - V_{dc})}\]  

\[(3.17)\]

where, \( I_{L0} = \frac{V_{dc} V_{th}}{L} \).

**Lamp peak current Boundary**  
As was mentioned before, there are two possible operating cases. The maximum lamp current \( \tilde{i}_{lp} \) of each case is different. When the peak current is reached after the gas breakdown, \( \tilde{i}_{lp} \) will be

\[Y_{lp} = (X_0 - X_{dc})\]

\[\tilde{i}_{lp} = (\tilde{V}_{lp} - V_{dc})\sqrt{\frac{C_d}{L}}\]  

\[(3.18)\]

Otherwise, when the peak current is reached before the gas breakdown, \( \tilde{i}_{lp} \) will be
3. Transformer-less Boost Resonant Converter

\[ Y_{lp} = (X_0 + X_{dc}) \]  
\[ \hat{i}_{lp} = (\hat{V}_{lp} + V_{dc}) \sqrt{\frac{C_{eq}^2}{L}} \]  

(3.19)

The boundary between both cases is obtained by analyzing the condition in which the maximum lamp current is achieved at the exact instance of the ignition. From the state planes is deduced that the boundary is given when \( X_R = X_{DC} \). Denormalizing the equation

\[ V_{dc} = -\hat{V}_{lp} + 2V_{th} \frac{C_g}{C_d} \]  

(3.20)

And, substituting \( \hat{V}_{lp} \) from Equation 3.17 and solving \( V_{dc} \), we can obtain the critical input voltage.

\[ V_{dc_{lim}} = \frac{V_{th}(C_d + C_g - \sqrt{(C_g + C_d)(\frac{V_{th}^2}{4L} + C_g)})}{C_d - \frac{V_{th}^2}{4L}} \]  

(3.21)

**Stable Operation Boundary** On the basis of the obtained expressions, it is possible to analyze and determine the stability conditions of this converter. The equation 3.17 shows a relation between the lamp voltage and the inverse of \((V_{th} - V_{dc})\). From this relation, it concludes that \( V_{dc} \) must be lower than \( V_{th} \), otherwise unstable operation occurs, since when \( V_{dc} \) approaches to \( V_{th} \), the lamp voltage will increase indefinitely.

\[ V_{dc} \leq V_{th} \]  

(3.22)

Figures 3.15 and 3.16 show the comparison between the state planes of a boost based converter in stable and unstable operation. Gray segments represent the changing stage (Subinterval A), blue ones are from the Discharge subinterval Before Ignition (B), and the orange segments are from the Discharge Subinterval - After Ignition (C).

![Figure 3.15: State plane of the Boost converter in stable operation](image)

![Figure 3.16: State plane of the Boost converter in unstable operation](image)
Lamp Power Throughout the state plane we can obtain an expression for the lamp average power $P_{lp}$ defined as,

$$P_{lp} = \frac{1}{T} \int_0^T v_{gas}i_{gas} dt = \frac{2V_{th}}{T} \int_{t_{br}}^{T/2} i_{lp} dt$$  \hspace{1cm} (3.23)

In the state plane of the subinterval C (Figures 3.13 and 3.14), it can be seen that once the gas has been broken down, the total voltage change in the dielectric is $\Delta X_d = X_0 - X_R$, and according to the capacitor voltage equation,

$$\Delta V_d = \frac{1}{C_d} \int_{t_{br}}^{T/2} i_{lp} dt \Rightarrow \int_{t_{br}}^{T/2} i_{lp} dt = C_d \Delta V_d$$  \hspace{1cm} (3.24)

Substitution of Equation 3.24 into Equation 3.23 and with the denormalization of $X_R$ (Equation 3.9) yields

$$P_{lp} = \frac{2V_{th} C_d \Delta V_d}{T}$$  \hspace{1cm} (3.25)

Finally, we can now substitute the peak lamp voltage (Equation 3.17) into Equation 3.25 to obtain

$$P_{lp} = fV_{th}(4V_{dc}C_gV_{th} + I_{L0}^2L) \frac{V_{th} - V_{dc}}{V_{th} - V_{dc}}$$  \hspace{1cm} (3.26)

Figure 3.17: Input Current waveform

Maximum Operation Frequency Figure 3.17 presents the current waveforms in half switching cycle. In order to guarantee a properly operation of the converter, the current $i_L$ must go down to zero before the next sequence starts. It means that the switching frequency must be lower than

$$f_{sw_{max}} = \frac{1}{2(t_{ch} + t_{pulse})}$$  \hspace{1cm} (3.27)

where, $t_{pulse} = t_{br} + t_{on}$. From the state planes and trigonometrical relationships the values of $t_{br}$ and $t_{on}$ can be found.
3. Transformer-less Boost Resonant Converter

\[ t_{on} = \sqrt{LC_d} \left( \pi - \arcsin \left( \frac{Y_R^*}{X_0 - X_{DC}} \right) \right) \]  
\[ t_{br} = \sqrt{LC_{eq}} \left( \arcsin \left( \frac{Y_R}{r_1} \right) - \arcsin \left( \frac{Y_{L0}}{r_1} \right) \right) \]  

3.1.2 Operating Point and Design procedure

From the lamp power equation can be seen that for a fixed lamp power, the needed peak lamp voltage can be reduced by increasing the switching frequency.

\[ P_{lp} = 4 f V_{th} C_d \left( \overline{V}_{lp} - \frac{V_{th} C_g}{C_{eq}} \right) \]  

(3.30)

Although this scenario can lead to the selection of a very high switching frequency, the speed of the switches will limit it. Furthermore, taking into account the results presented by [4] about the impact on the UV production of the frequency \( f_{lp} \), current intensity \( J \) and duty cycle \( D \), the following considerations are analysed to accurately select the operation point[4]:

- **Impact of \( J \) over UV Production**: With current intensity above 128 mA, an increment in the lamp UV production of about 25 % for the same electrical power injected into the lamp, is obtained.
- **Impact of \( f_{lp} \) over UV Production**: In Figure 3.18 it is evaluated the UV production keeping a constant \( J \) and varying \( f_{lp} \) [4]. The graphics show that the smaller current intensity, the higher the impact of the frequency on the lamp performance. And for high current intensity, between 50 kHz and around 140 kHz the impact of the frequency is negligible and the UV output increases linearly with \( P_{lp} \).

![Figure 3.18](image.png)

**Figure 3.18**: Impact of the operating frequency in the UV output for \( J=250 \) mA (left) and for \( J=128 \) mA (right) [4]

Considering the aforementioned aspects, a switching frequency of \( f_{lp} = 150 \) kHz was selected. According to equation 3.30, for a lamp power of \( P_{lp} = 100W \), a peak lamp voltage of 3250 V is needed.
In order to easily select the components an spreadsheet with all previous formulas was designed (Figure 3.19). The inputs of the spreadsheet are: Lamp parameters \((V_{th}, C_g, C_d, P_{lp})\), switching frequency \(f\), and the relation between the \(T_{on}\) and half cycle \(T_{sw}/2\). A sweep through different input voltages \(V_{dc}\) was done, highlighting when the requested frequency is higher than the maximum operation frequency, and also ensuring that the peak current happen after the gas breakdown.

![Figure 3.19: Screenshot spreadsheet tool](image)

### 3.2 High Voltage SiC Switches Proposal

Connecting a chain of normally-on JFETs with a low voltage MOSFET in a Super Cascode configuration (Figure 3.20) it is possible to easily control them using only one gate driver per Super Cascode chain [24] [22]. The more the high voltage devices with a blocking voltage \(V_{BR}\) are stacked, the higher the possible operating voltage of the resulting switch, \(V_{BR_{total}} = n \times V_{BR}\) where \(n\) is the number of single switches.

![Figure 3.20: Super Cascode schematic](image)
3.2.1 Working Principle

**Turn On Process** When a positive gate voltage is applied to the MOSFET, the source and gate of J1 are shorted then it will conduct. $v_{gs,J1}=0$

As the JFETs are normally on devices, the next JFET is conducting as well, since the source of J2 is connected to the anode of D1 via the turned on J1 and MOSFET. The cathode voltage of D1 can not be less than the diode forward voltage $V_F$, so the gate voltage of J2 is higher than $-V_F$, which is above the threshold voltage of the JFETS ($V_{th} \approx -7$ V). Similarly, J3 is turned on sequentially following the abovementioned process.

$$v_{gs,J2} > -V_{F,D1} > V_{th,J2}$$

In the real circuit, the gate voltage of J2 is between $-V_F$ and the forward voltage of the gate diode of the JFET, depending on the leakage current distribution in the JFETs and diodes [25].

**Turn Off Process** In order to turn off the switch, the MOSFET must be turned off and consequently, the drain voltage of the MOSFET increases until the pinch-off voltage between the gate and source of J1 is reached.

$$v_{gs,J1} = -v_{ds,M} = V_{th,J1}$$

At this point $v_{ds,M}$ is fixed and J1 turns off and blocks the rising drain–source voltage until the avalanche breakdown of D1 occurs. By reaching the avalanche of D1, the gate voltage of J2 stays fixed but the source of J2 continues increasing until the gate - source voltage of J2 is equal to $-V_{th}$

$$v_{gs,J2} = -v_{ds,M} - V_{ds,J1} + V_{br,D1} = V_{th,J1}$$

If the pinch-off voltages of the JFETs are equal, we can guarantee that the maximum voltage of J1 will be

$$v_{ds,J1} = -(V_{th,J1} - V_{th,J2} - V_{br,D1})$$

$$v_{ds,J1} = V_{br,D1}$$

After the turnoff process, the static voltage distribution is determined by the avalanche voltage of diodes D1 and D2. For a controlled and stable avalanche, a certain leakage current through the diodes is required. In order to guarantee this leakage current, resistors (R1 and R2) must be connected between the gate and the source of the JFETs (Figure 3.21) [24]. In such a way, the leakage current is defined by the resistance and $v_{gs,J}$ (it means $V_{th}$ in the off state).

3.2.2 Transient Behavior

In order to avoid over-voltages, capacitors CD and resistors RD are included for damping oscillations during the switching transients, as is shown in the Figure 3.21 [23].
A larger capacitance value for CD results in more synchronous switching transients. Starting from the OFF state all the capacitors are equally charged, in the turn-on process, as soon as J1 starts to conduct, the potential of the source of J2 decreases, however due to $C_{D1}$ the potential of the J2 gate is fixed for a limited time, so that the gate voltage of J2 starts to increase as soon as the potential of the source starts to decrease [23].

Nevertheless, at turnoff, a too large value for CD results in a more synchronous switching operation but an unbalanced voltage distribution since, at the beginning of the turnoff, the capacitors are discharged, so that they hold the gate potential of J2 down. When J1 now starts to turn off, the gate voltage of J2 immediately becomes negative and turns off J2 faster than J1, so that J2 is blocking the largest share of the voltage [23].

Figure 3.21: Complete Super Cascode schematic - Leakage resistances (R1 and R2), and dynamically balancing (RD1, RD2, CD1 and CD2).

Figure 3.22: LTSPICE Super Cascode Simulation 3000V - 6A
In Figure 3.22, the super cascode switch is simulated in LTSPICE with the models of the JFETS, the value of CD was selected experimentally to guaranty a fast turn off with a good voltage distribution.

### 3.3 Hardware Implementation

**SiC Power Semiconductors** According to the current-voltage characteristics of the switches, the devices need to have a high frequency and high voltage thyristor-like behavior. Figure 3.23 presents the structure of the implemented switch using the JFET Super Cascode.

The selected JFET provided by USCi. UJN1208K is a 1200 V SiC Normally-On JFET, with a on resistances of $R_{DS(ON)} = 80 \, \text{m} \Omega$, and an intrinsic capacitance of $C_{oss} = 94 \, \text{pF}$. The low voltage MOSFET is the STD75N3LLH6.

The series connection of the diodes is made using a C3D10170H fabricated by CREE, it is a SiC 1700 V Shottky Rectifier, with Zero Reverse Recovery Current and Zero Forward Recovery Voltage.

Diodes connected in series tend to unequally share the voltage across the chain in blocking conditions because of the variations in reverse characteristics: leakage currents and turn-off switching parameters [26]. In order to equalize the voltage, a resistor is connected in parallel of each diode, the calculation of the sharing resistors was made according to [26].

**Power Switch Gate Drivers** The PWM gate signals are generated by the FPGA DE0-nano. In order to change the parameters of the boost converter without reprogramming the FPGA, this is communicated with a Python program through the USB port the computer. The parameters that can be modified are: charging time $t_{ch}$, switching frequency, dead time and overlapping time.

---

**Figure 3.23:** Thyristor structure

**Figure 3.24:** Circuit Diagram for Fiber Optics isolation [27]
In high-voltage systems, an optical fiber connection is usually used for transmitting the control signals. This isolation method has clear advantages compared to other technologies because theoretically has an infinite isolation, a very high EMI immunity, no coupling capacity and unlike pulse transformers it can transfer AC and DC signals [27]. Figure 3.24 shows the used configuration of transmitter, receiver and fiber optic link manufactured by Avago Technologies.

Each driver is powered by a 15 V voltage power supply with an isolation of 5.2 kVDC. The fiber optics signal is transmitted and received by the high speed gate driver IXDD609 which control the low voltage MOSFET in the entrance of the Super Cascode Chain. The implemented prototype and the schematic are shown in the Figure 3.25 and 3.26 respectively.

3.4 Simulation and Experimental Results

Simulation Results  Figure 3.27 presents the output waveforms of the converter simulated with the components values of Table 3.1. This simulation does not consider any parasitic elements. The output voltage in the simulation is around 3.4 kV for a power in the lamp of 98 W. This demonstrates that the design procedure is sufficiently precise for the ideal converter sizing.

With the aim of verifying the behaviour of the converter with the complete model of the components, the circuit has been simulated in LTSPICE using the models provided by the manufacturer.
Figure 3.28. presents the simulation results. In the top figure are the input current and the currents flowing through S1 and S3, in this figure can be seen that although in the first semicycle S3 is turned off, the current input is almost equally derived for both legs of the bridge. The main reason of this behaviour is that the output capacitance of the JFETs are similar to that of the DBD lamp capacitance. Once the voltage on the gas reaches the threshold voltage, as the equivalent capacitance of the lamp changes, it is seen that the current distribution varies, due to a variation in the capacitive current divider.

Table 3.1: Components Values. Transformer-less

<table>
<thead>
<tr>
<th>Component</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L$</td>
<td>1.93 mH</td>
</tr>
<tr>
<td>$V_{in}$</td>
<td>550 V</td>
</tr>
<tr>
<td>$t_{ch}$</td>
<td>1.4 µs</td>
</tr>
<tr>
<td>$f$</td>
<td>150000 Hz</td>
</tr>
</tbody>
</table>

Figure 3.27: Ideal Simulation Results

Figure 3.28: LTSPICE Simulation
3. Transformer-less Boost Resonant Converter

Considering the complete model of the switches, the maximum voltage in the lamp is around 2000 V while in the ideal simulation the lamp voltage was 3400 V for 100 W. Under this scenario, it must not be possible for the lamp to be turned on, and the efficiency of the system will be extremely low as will be shown below.

**Experimental Results**  Figure 3.30 presents the experimental results of the converter with an input voltage of 150 V and Figure 3.29 shows the simulation using Spice models. We can clearly see that the maximum lamp voltage in both cases is almost 600 V but ideally the lamp voltage had to be 1800 V. In Figures 3.31 and 3.32, we present the same results for an input voltage of 250 V.

![Figure 3.29: Simulation LTSPICE Vin=150 V](image1)

![Figure 3.30: Experimental Results Vin=150 V](image2)

![Figure 3.31: Simulation LTSPICE Vin=250 V](image3)

![Figure 3.32: Experimental Results Vin=250 V](image4)

Even if the input voltage is increases until the 550 V, where the operating point is expected, the lamp will remain off. Due to the high losses of the converter, in order to operate the lamp at 100 W, the input voltage needs to be beyond the maximum ratings of the available equipment and the dissipation heat of the devices would cause their damage. As this project intends to improve the efficiency, the next chapter presents an explanation of the malfunctioning of the converter for the purpose of design a better power supply.
3. Transformer-less Boost Resonant Converter
4 Optimum Transformer Turns Ratio

After the implementation of the transformer-less power supply, it was evident that the behaviour of the converter is far from the ideal operation. In spite of the elimination of the parasitic elements associated with the transformer, since the capacitances of the switches $C_{sw}$ are comparable to the equivalent capacitance of the lamp the waveforms are affected impeding the lamp ignition.

As well as the parasitic capacitance of the transformer $C_{p}$, the capacitances of the switches $C_{sw}$ appear in parallel to the lamp.

When a transformer with a high turns ratio is used, the capacitances $C_{sw}$ are reflected in the secondary side with a much lower value ($C_{sw/sec} = C_{sw}/n^2$) therefore they can be neglected and the influence of the transformer capacitance will be more relevant.

As the transformer turns ratio decreases, the capacitance $C_{p}$ can be decreased as well, while the reflected $C_{sw}$ increases so its effects are more and more visible.

In order to have a better understating of the effects of the parasitic elements and explain the malfunctioning of the transformer-less power supply, in the following section, the analysis of the Boost Based converter with the capacitances of the switches is presented.

4.1 State plane analysis with parasitic capacitances

By including the capacitances $C_{sw}$ in the analysis of the Boost Based Converter, new subintervals of commutation must be described. Figures 4.1 - 4.4 show the first four phases, in the fifth phase all the switches are turned off, followed by the negative semi-cycle. Only the intervals corresponding to the positive current are analyzed, and due to the symmetry of the waveforms, the negative semi-cycle can be deduced from the positive.

Figure 4.1: $t_o < t < t_{ch}$
Figure 4.2: $t_{ch} < t < t_1$
Figure 4.3: $t_1 < t < t_{br}$
Figure 4.4: $t_{br} < t < t_{off}$
Figure 4.5 shows the waveforms for the currents and voltages of the converter when the parasitic capacitance of the switches is comparable with the lamp capacitance.

**Figure 4.5:** Waveforms for Boost - Based Converter with strong effects of $C_{sw}$

**Subinterval** $t_o < t < t_{ch}$: Figure 4.1. When S1 and S2 are turned-on the current $i_{L1}$ starts to ramp up, at $t_{ch}$ the current is given by Equation 3.1. In this phase the lamp voltage remains negative and constant, consequently, diodes D3 and D4 are off and there are no effects of $C_{sw}$ in this phase.

**Subinterval** $t_{ch} < t < t_1$: Figure 4.2. when switch S4 is turned on and switch S3 is turned off, an LC series resonant circuit is obtained between L and the equivalent lamp capacitance.

$$C_{eq1} = \frac{C_d C_g}{C_d + C_g}$$

Due to the resonance, the lamp current grows, and the gas voltage, $v_{gas}$, increases together with the lamp voltage $v_{lp}$. If $v_{lp}$ is negative, the diodes D2 and D3 are off, when $v_{lp}$ crosses through 0, the diodes are turned on and the next subinterval of commutation begins.
Figure 4.6: Equivalent Circuit Subinterval $t_{ch} < t < t_1$

Figure 4.7: State Plane Subinterval $t_{ch} < t < t_1$

Figure 4.8: Equivalent Circuit Subinterval $t_1 < t < t_{br}$

Figure 4.9

Figure 4.10

4. Optimum Transformer Turns Ratio

Figure 4.6: Equivalent Circuit Subinterval $t_{ch} < t < t_1$

Figure 4.7: State Plane Subinterval $t_{ch} < t < t_1$:

Figure 4.3.

Once the diodes D2 and D3 are turned on, the capacitances of the switches S2 and S3 are connected in parallel to the lamp. The current divider between $2C_{sw}$ and $C_{eq1}$ defines how much current will flow through the lamp.

\[ i_{lp} = i_L \frac{C_{eq1}}{2C_{sw} + C_{eq1}} \]  

(4.3)

In the Figure 4.8 is the equivalent circuit of the subinterval where:

\[ C_{eq2} = \frac{C_{eq1}2C_{sw}}{C_{eq1} + 2C_{sw}} \]

In Figures 4.9 and 4.10 the state plane analysis for the case A (lamp peak current after the breakdown) and case B (lamp peak current before the breakdown) are shown.

\[ Y_1^2 = (X_{DC} - Xo)^2 + Y_{LO}^2 - X_{DC}^2 \]

(4.1)

\[ r_1^2 = (X_{DC} - Xo)^2 + Y_{LO}^2 \]

(4.2)

Subinterval $t_1 < t < t_{br}$: Figure 4.3.

Figure 4.9 and 4.10 the state plane analysis for the case A (lamp peak current after the breakdown) and case B (lamp peak current before the breakdown) are shown.
From the trigonometrical relations it can be obtained the breakdown current.

\[ Y_R^*^2 = r_2^2 - (X_{DC} - X_R)^2 \]  \hspace{1cm} (4.4)

where, \( X_R \) is given by equation 3.9 and \( r_2 \) is

\[ r_2^2 = X_{DC}^2 + Y_1^*^2 r_2^2 = X_{DC}^2 + \left[ (X_{DC} - X_o)^2 + Y_{LO}^2 - X_{DC}^2 \right] \frac{C_{eq1}}{C_{eq2}} \]

Subinterval \( t_{br} < t < t_{off} \): Figure 4.4. When the ignition voltage is reached, the lamp behaves as a capacitor in series with a \( V_{th} \) source voltage. As is shown in the Figure 4.11, the circuit is reduced to the connection of an equivalent capacitance in series with an equivalent voltage source of values:

\[ V_{eq} = V_{th} \frac{C_d}{C_d + 2C_{sw}} \]

\[ C_{eq3} = C_d + 2C_{sw} \]

In this case the current divider determines how much current is available to hold up the discharge and to transfer power into the gas.

\[ i_{lp} = i_L \frac{C_d}{2C_{sw} + C_d} \]  \hspace{1cm} (4.5)

From the state planes of Figures 4.12 and 4.13, the breakdown current can be expressed as:

\[ Y_{R**}^*^2 = (X_o - X_{DC})^2 - (X_{DC} - X_R)^2 \]  \hspace{1cm} (4.6)
4. Optimum Transformer Turns Ratio

Figure 4.12: State plane Subinterval $t_{br} < t < t_{off}$ in Operating Case A.

Figure 4.13: State plane Subinterval $t_{br} < t < t_{off}$ in Operating Case B.

An expression for the normalized lamp peak voltage can be obtained by relating the normalized current at the breakdown from the state plane before the ignition (Equation 4.4) with the state plane after the ignition (Equation 4.6).

\[
X_o = \frac{\sqrt{C_{eq1}^4(X_{DC}^2 - Y_{LO}^2)} + C_{eq1}^3(4X_{DC}^2C_{eq3} - C_{eq2}(2X_{DC}^2 - Y_{LO}^2)) + \sqrt{C_{eq1}(C_{eq1} - C_{eq2})}}{X_{DC}^2C_{eq2}(C_{eq2}^2 - 4C_{eq2}C_{eq3} + 4C_{eq3}^2) + 4C_gX_{th}C_{eq1}(C_{eq2} - C_{eq3})(2X_{DC}C_{eq3} + C_gX_{th}) + \sqrt{C_{eq1}(C_{eq1} - C_{eq2})}} - 4C_g^2X_{th}^2C_{eq3}(C_{eq2} - C_{eq3})) - X_{DC}C_{eq1}^2 + X_{DC}C_{eq1}(C_{eq2} - 2C_{eq3}) - 2C_gX_{th}(C_{eq2} - C_{eq3})}{C_{eq1}(C_{eq1} - C_{eq2})} \tag{4.7}
\]

The limit between the operating case A and B can be obtained solving $X_{DC}$ from the condition $X_R < X_{DC}$.

\[-X_o + 2X_{th}\frac{C_g}{C_{eq1}} < X_{DC} \tag{4.8}\]

Replacing 4.7 in 4.8, and solving for $X_{DC_{\text{min}}}$,

\[X_{DC_{\text{min}}} = \frac{2C_g\sqrt{LX_{th}}(2\sqrt{LC_{eq3}} - \sqrt{LC_{eq1}^2 - C_{eq1}(LC_{eq2} - 3LC_{eq3} - t_{ch}^2) + C_{eq3}(LC_{eq2} + t_{ch}^2)})}{C_{eq1}(LC_{eq1} - LC_{eq2} + 4LC_{eq3} - t_{ch}^2)} \tag{4.9}\]

Finally, the maximum operating frequency is

\[f_{lp_{\text{max}}} = \frac{1}{2(t_{ch} + t_1 + t_{br} + t_{on})} \tag{4.10}\]

From the figures 4.7 4.9 and 4.12:

\[t_1 = \sqrt{LC_{eq1}}\left[tan^{-1}\left(\frac{Y_1}{X_{DC}}\right) - tan^{-1}\left(\frac{Y_{LO}}{X_{DC} + X_{o}}\right)\right] \tag{4.11}\]
Table 4.1: Comparison of the Operating Point Transformer-less Boost Based Converter

<table>
<thead>
<tr>
<th></th>
<th>V_{lp}</th>
<th>i_L</th>
<th>t_{on}</th>
<th>P_{lp}</th>
</tr>
</thead>
<tbody>
<tr>
<td>Ideal</td>
<td>3247 V</td>
<td>0.57 A</td>
<td>666 ns</td>
<td>100 W</td>
</tr>
<tr>
<td>Theoretical with (C_{sw})</td>
<td>2122 V</td>
<td>0.49 A</td>
<td>687 ns</td>
<td>22.6 W</td>
</tr>
<tr>
<td>LTSPICE Simulation</td>
<td>2112 V</td>
<td>0.53 A</td>
<td>670 ns</td>
<td>25.62 W</td>
</tr>
</tbody>
</table>

\[
t_{br} = \sqrt{L_{eq2}} \left[ \cos^{-1}\left(\frac{X_{DC} - X_R}{r_2}\right) - \cos^{-1}\left(\frac{X_{DC}}{r_2}\right) \right] \quad (4.12)
\]

\[
t_{on} = \sqrt{L_{eq3}} \left[ \pi - \sin^{-1}\left(\frac{Y_{R}^{**}}{r_3}\right) \right] \quad (4.13)
\]

Due to the complexity of the previously found formulas, the excel spreadsheet was used for their computation. Figure 4.14 presents the comparison between the ideal analysis and the analysis with parasitic effects for the converter implemented in the transformer-less power supply (Components values of Table 3.1). In Table 4.1 the theoretical operating point with \(C_{sw} = 94\) pF is compared with the ideal operation and the LTSPICE simulation (Figure 3.28). As can be seen, the elaborated analysis is a good approximation for the proper understanding of the converter.

**Figure 4.14**: Current waveforms for the Transformer-less Boost Based Converter

### 4.2 Losses Analysis

The most significant losses in the converter are produced by the conduction and commutation, therefore they will be analysed neglecting the gate charge losses and the losses of the body diode conduction and body diode reverse recovery as follow:

#### 4.2.1 Switching Losses

Switches S1 and S3 have the same voltage and current waveforms, the only difference is that they are shifted T/2, so in terms on the switching losses it is enough to only analyse one\(^7\). In the same manner S2 and S4 can be analysed together.

In the first commutation interval, S1 is turned on under ZCS due to the discontinuous conduction mode. Then, S1 has a forced turn off and commutation losses
4. Optimum Transformer Turns Ratio

associated to the stored energy in the drain-source capacitance $C_{sw}$. On the other hand, S3 is turned on after the charging time with a non-zero current.

Since before turn-on, the energy stored in the capacitor $C_{sw}$ is $E_{Csw} = \frac{1}{2}CV^2$, the power losses can be calculated as:

$$P_{on} = \frac{1}{2}f_{sw}CV^2$$  \hspace{1cm} (4.14)

**Conduction Losses** Conduction losses in the MOSFET can be calculated using the drain-source on-state resistance (RDSon):

$$P_c = \frac{1}{T_{sw}} \int_0^{T_{sw}} p(t)dt = R_{DSon}I_{rms}^2$$  \hspace{1cm} (4.15)

### 4.3 Selection of the Optimum turns ratio

In order to know what is the optimum transformer turns ratio (N) with which the highest efficiency can be achieved, the power losses are studied looking for a better appreciation of the effects of N in the behaviour of the converter.

In general, at lower turns ratios the voltages in the switches are higher and the effects of the capacitor $C_{sw}$ are stronger. The higher N, the lower the effects of $C_{sw}$ and its losses, but as the current in primary side increases linearly with N, the conducting losses will increase as well, without mentioning that the parasitic capacitance of the transformer rises.

To be able to compare the efficiency of the power supply at different turns ratios, the theoretical design of 10 converters (from N=1 to N=10) was realized, guaranteeing that under the effects of $C_{sw}$, the lamp power, the $t_{on}$ time and the switching frequency were the same. Following the same considerations presented in the Section 3.1.2, the selected operating point is:

<table>
<thead>
<tr>
<th>$f_{sw}$</th>
<th>$P_{lp}$</th>
<th>$V_{lp}$</th>
<th>$%t_{on} \equiv D$</th>
<th>$t_{on}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>150 kHz</td>
<td>100 W</td>
<td>3247 V</td>
<td>20 %</td>
<td>666 ns</td>
</tr>
</tbody>
</table>

Although the parasitic capacitance of the transformer can be predicted by means of analytical expressions as is presented in [8], this method only provides an efficient tool for high step-up ratios. In that regard, the value of $C_p$ can not be easily formulated as a valid function for the whole N sweep. Therefore, the study of the efficiency will not consider the effects of $C_p$.

As was mentioned before, the major losses are given by the conduction and the energy stored in $C_{sw}$.

The conduction losses are divided between the switches, the diodes and the input inductance $L$. For each case, the RMS current is numerically estimated in excel with n=2000 samples per period. It is worth to mention that the series resistances were not used in the theoretical analysis and they are only utilized for the losses estimation.
4. Optimum Transformer Turns Ratio

The inductance losses are function of the RMS current of $i_L$ and the series resistance $R_L$ that can be calculated with the coil length and the wire resistivity $\rho$.

$$P_L = R_L \sqrt{\frac{1}{n} \sum_{i=1}^{n} i_i^2}$$  \hspace{1cm} (4.16)

The conduction losses of the switch are calculated with the drain-source on-state resistance ($R_{DSon}$) and the rms current of each switch.

$$P_{sw1} = R_{DSon} \sqrt{\frac{1}{n} \sum_{i=1}^{n} i_{sw1i}^2}$$  \hspace{1cm} (4.17)

The conduction losses in a diode appear when the diode is in forward conduction mode due to the on-state voltage drop ($V_F$).

$$P_{D1} = V_F \sqrt{\frac{1}{n} \sum_{i=1}^{n} i_{sw1i}^2}$$  \hspace{1cm} (4.18)

And finally, the losses relating the energy stored in the capacitor $C_{sw}$ are given by equation 4.14.

$$P_{C_{sw}} = \frac{1}{2} f_{sw} C_{sw} V_{lp}^2$$  \hspace{1cm} (4.19)

In Figures 4.15 and 4.16, the results of the efficiency comparison between converters with different transformer turns ratio is presented. Additionally, Figures 4.17, 4.18, 4.19 and 4.20 show the current waveforms of each converter.

| CONVERTER NUMBER | OPERATING POINT | L (\mu H) | f (kHz) | RMS CURRENT | ON STATE | CONDUCTANCE | EFFICIENCY | TOTAL CONDUCTANCE | CONDUCTOR LOSSES | ON STATE LOSSES | ENERGY LOSSES | TOTAL LOSSES | EFFICIENCY |
|------------------|-----------------|---------|--------|-------------|---------|-------------|-----------|---------------------|-----------------|---------------|--------------|-------------|-----------|-----------|
| 1                | 950             | 0.000707 | 6.546    | 6.347       | 100     | 6.437       | 0.380     | 0.18        | 0.18            | 0.16         | 0.28         | 0.28        | 0.30       | 100.42    |
| 2                | 460             | 0.000304 | 6.415    | 6.37       | 100     | 6.511       | 0.358     | 0.15        | 0.15            | 0.10         | 0.23         | 0.23        | 0.25       | 42.86     |
| 3                | 340             | 0.000256 | 5.117    | 5.117       | 100     | 5.116       | 0.253     | 0.12        | 0.12            | 0.10         | 0.20         | 0.20        | 0.20       | 22.10     |
| 4                | 220             | 0.000340 | 5.216    | 5.216       | 100     | 5.217       | 0.280     | 0.13        | 0.13            | 0.11         | 0.22         | 0.22        | 0.22       | 43.05     |
| 5                | 150             | 0.000952 | 6.164    | 6.164       | 100     | 6.167       | 0.333     | 0.19        | 0.19            | 0.15         | 0.30         | 0.30        | 0.30       | 15.57     |
| 6                | 340             | 0.000254 | 5.117    | 5.117       | 100     | 5.116       | 0.253     | 0.12        | 0.12            | 0.10         | 0.20         | 0.20        | 0.20       | 22.10     |
| 7                | 120             | 0.000329 | 5.468    | 5.468       | 100     | 5.468       | 0.323     | 0.17        | 0.17            | 0.17         | 0.33         | 0.33        | 0.33       | 14.18     |
| 8                | 105             | 0.000331 | 5.116    | 5.116       | 100     | 5.117       | 0.253     | 0.12        | 0.12            | 0.10         | 0.20         | 0.20        | 0.20       | 22.10     |
| 9                | 90              | 0.000320 | 5.314    | 5.314       | 100     | 5.315       | 0.305     | 0.20        | 0.20            | 0.18         | 0.32         | 0.32        | 0.32       | 18.77     |
| 10               | 80              | 0.000347 | 5.417    | 5.417       | 100     | 5.419       | 0.323     | 0.17        | 0.17            | 0.17         | 0.33         | 0.33        | 0.33       | 14.18     |

Figure 4.15: Efficiency Results varying N with the same operating point - $C_{sw} = 50$ pF, $R_{DSon} = 160$ m\(\Omega\), $V_F = 1.7$ V and $\rho = 0.013 \, \Omega \, m$.

Figure 4.16: Efficiency Results varying N with the same operating point.
According to these results, the maximum theoretical efficiency is 88% with a turns relation of N=6. The next section shows the implementation of this converter, the values of the components are in Table 4.3.

### Table 4.3: Components Values of the Power Supply with Transformer

<table>
<thead>
<tr>
<th>Component</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>N</td>
<td>6</td>
</tr>
<tr>
<td>L</td>
<td>44 µH</td>
</tr>
<tr>
<td>Vin</td>
<td>140 V</td>
</tr>
<tr>
<td>t_{ch}</td>
<td>581 ns</td>
</tr>
<tr>
<td>f</td>
<td>150000 Hz</td>
</tr>
</tbody>
</table>

### 4.4 Hardware Implementation

#### 4.4.1 High Frequency SiC Thyristor

As was mentioned before, due to the current-voltage characteristics of the switches, it is necessary to use switch capable of blocking bipolar voltages. The high frequency thyristor-like behavior switch provides this characteristics and can be implemented with a MOSFET in series with a Diode.

The advantages of SiC MOSFETs have been presented extensively in the literature [28] [29], one of the key advantages is the high temperature capability afforded
by the wide bandgap, besides SiC MOSFETs reduces switches losses compared with Si MOSFETs, and have a higher breakdown electric field strength [30]. These properties allow a high power device to block high voltages in the blocking mode [31].

The selected MOSFET is a SiC C2M0160120D by CREE, it is a 1200 V device, with a $R_{DS_{ON}} = 160 \, m\Omega$, and a $C_{sw} = 47 \, pF$. And the selected Diode is the C3D10170H also fabricated by CREE.

### 4.4.2 SiC MOSFET Driver Module

Since, by using the transformer the voltages in the switches is decreased, the use of fiber optics for the signal isolation will not be necessary and a cheaper solution can be implemented.

The considerations presented in [32] were used for the design of the driver of the switches. Each switch was implemented in a module which consists of an isolated DC-DC converter, and opto-isolator, and the gate IC-driver. Figure 4.21 shows the module.

Figures 4.22 and 4.23 show the prototype and its respective schematic.
4.5 Simulation and Experimental Results

Taking into account the spice models provided by CREE, the converter was simulated in LTSPICE. The results of the simulation are shown in the Figure 4.24. The experimental results are presented in Figure 4.25.
The measured efficiency of the converter is 84%. As it is less than the expected efficiency, a converter with a turns ratio of 4.5 was implemented to have a reference of comparison. Figure 4.26 shows the results.

The converter with a turns ratio of N=4.5 obtained an efficiency of 87%. Although the losses analyses showed that the higher efficiency will be achieved at N=6, it must be considered that the parasitic capacitance of the transformer was not considered in the analysis.
5 Conclusions

- In order to select a topology for the implementation of the power supply, the review and comparison of the Boost Based Converter, the Series Resonant Inverter and the Buck-Boost converter was realized. Due to the low input voltage requirement, the Boost Based converter was selected and analyzed.
- By removing the transformer the blocking voltage of the switches must be increased. The Super Cascode switch provides a solution by staking SiC JFETs.
- The experimental results of the transformer-less power supply showed evident differences between the implementation and the ideal operation. As the capacitances of the switches are comparable to the lamp equivalent capacitance, the ignition of the lamp was not achieved in the transformer-less power supply due to the high losses.
- The power supply without transformer can be useful for other type of lamps with higher equivalent capacitances. Further applications of the transformer-less converter using Super Cascode switches may include solar inverters and DC-DC High voltage converters.
- Throughout the state plane analysis it is possible to dimensioning the components values for the Boost based converter considering the impact of the parasitic capacitance of the switches. In order to ensure that the theoretical analysis properly describes the effects of $C_{sw}$, the designed converters were verified in simulation and experimentally.
- The losses of the different components were described and studied looking for an estimated efficiency. A sweep through twelve different cases of the converter varying the transformer turns ratio from $N=1$ to $N=20$ and ensuring that the lamp operates at the same operating point was done. The losses distribution and efficiency of all the cases were compared to select the optimum case and the maximum theoretical efficiency was obtained with $N=6$ and 88%.
- The worst theoretical efficiency was obtained in the transformer-less case (38%), in which the losses overcome 150 W. When there is no transformer, the capacitances of the switches are directly connected in parallel to the lamp, therefore the energy stored in $C_{sw}$ and the loss power associated with it are the highest.
- On the other hand, higher turns ratios implies lower losses of $C_{sw}$ but higher conduction losses, consequently in order to obtain the highest efficiency, the optimum turns ratio must compensate both type of losses.
- Comparing with the previous work [4], the efficiency of the system was improved. In the previous converter with $P=100$ W, $N=10$ $f=80$ kHz and Si devices the efficiency is registered about 80 % and with SiC devices, $P=100$
5. Conclusions

W, N=4.5 and f=150 kHz the efficiency is 87%.


[32] CREE. Sic mosfet isolated gate driver. Application Note CPWR-AN10, REV -C.