Design of an on-board charger for plug-in hybrid electrical vehicle (PHEV)

Master of Science Thesis 2009

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Abstract

The purpose of this project is to design an on-board charger with the aim of charging a plug-in vehicle battery. This charger is able to control the values of the load voltage and current and then, maintain them at a desirable value. A first converter is used to transform the grid 50Hz electrical quantities into DC quantities. A second converter adjusts the levels to the values required by the battery and moreover, provides a galvanic isolation. The control of the first converter is realized by using a power factor corrector function. The control of the second converter allows supplying the battery with correct voltage and current values. This method is one of the most efficient to design an electronic supply with a low current harmonic impact to the grid.

This converter topology is investigated in this report. The choice of the structure and the choice of the components are explained, the magnetic components, inductors and transformer, are designed. A simulation, realized in Matlab Simulink, allows the understanding of the working principle of the power factor corrector and the full bridge converter. Simulations allow checking the theory part and the determination of the problems occurring in the on-board charger. The efficiency of the on-board charger is also determined with Matlab. Determination of the loading cost and loading time needed is investigated.

Acknowledgment

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Contents

1) Introduction

Background

The oil reserves are not inexhaustible, it is therefore necessary to find additional energy sources. This change has also become necessary due to the fact that the burning of oil gives a negative environmental impact. Nowadays, most cars are using internal combustion engines but car manufacturers are considering using different kinds of energy sources that reduce the pollution. The energy carrier which seems be the most usable is the electrical energy. A hybrid vehicle use two propulsion sources, usually one is an electric motor and the other is an internal combustion engine, in addition different technologies exist. The two base types are the parallel hybrid vehicle, where the movement of both propulsion sources are added, and the series hybrid vehicle, where the internal combustion engine powers the electric motor. These two kinds of hybrid vehicle can go solely on the electrical source for a very limited range. A plug-in hybrid electrical vehicle (PHEV) is a parallel/series hybrid vehicle which can be reloaded by a connection to an electrical network.

In order to make this charging easy, an idea is to have an on-board charger to charge the plug-in hybrid vehicles. This supply should to be able to plug into a classical socket for simplicity reasons and moreover, it should to be a grid friendly charger in order not to pollute the electrical network.

Research is already being done on these on-board chargers, and one important objective is to increase the energy efficiency and further reduce the power quality impact. This charger can also be made bidirectional in order to be useful as a power supply or be connected to the grid to give power.

Purpose

The purpose of this project is to design an on-board charger after having compared two different technologies. This charger should be energy efficient, have a low weigh and have a low volume.

Layout of thesis

First, the theory study about the power factor corrector, the full bridge converter and the battery with the aim of understanding the operation and design of the on-board charger. Then, simulations in simulink of the power factor corrector and the full bridge to checking the theoretical study. After, will be the choice of the components and design of the magnetic components will be performed. Finally, it will be the efficiency and the on board charger and the practical design the transformer will be presented.

2) Theory

The theoretical study is about two different structures for an on board charger. The first structure which is studied has a unidirectional power transfer and the second one has a bidirectional power transfer. This study presents the operation of the on board charger (OBC), more particularly the different converters (PFC, inverter, transformer, DC/DC converter) and the control of these converters. All these parts are detailed below in different chapter.

1) Unidirectional on board charger

This converter makes energy transfer possible only in one way, from the grid to the battery. Today, this kind of structure is the most far used in charger applications.

Figure 1 shows a general diagram of the on board charger with the different converters, and also with the load (battery) and the grid.

Figure 2 shows the power part of the on board charger different converter

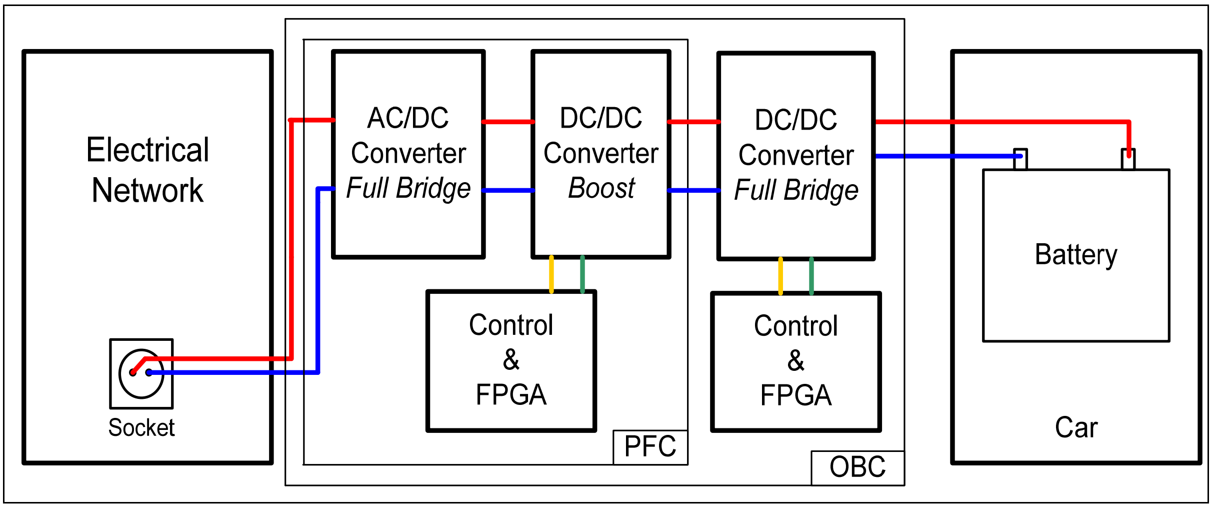


Figure 1 General diagram of the on-board charger

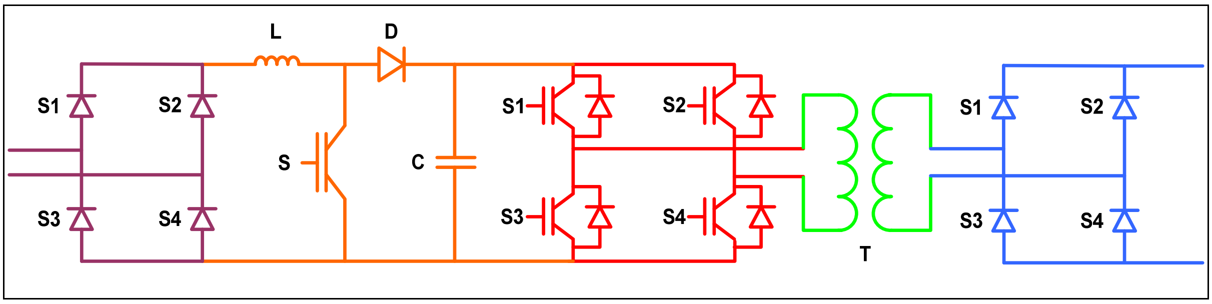


Figure 2 Power part of the on board charger

A) PFC

As illustrated in the previous diagram, a PFC is an AC to DC converter that is made up by an AC to DC converter and a DC to DC converter (boost converter). The aim of a PFC is to take a current close to a sinusoidal waveform from the network. The input current of a PFC is in phase with the grid voltage. An important feature of the PFC control is a voltage sensor which measures the load voltage and also a current sensor which measures the inductor current in the boost converter. Using this information, the PFC control part can be realized.

These parts are more detailed below.

AC/DC Stage

A rectifier converts a sinusoidal voltage to a DC voltage. There are different ways to make this transformation, using diodes, or thyristors or diodes and thyristors. When it is not necessary to control the output voltage value, the use of diode is the easiest method. This circuit is cheap and easy to design and doesn't need to be controlled. Figure 3 shows this circuit with diode usage.

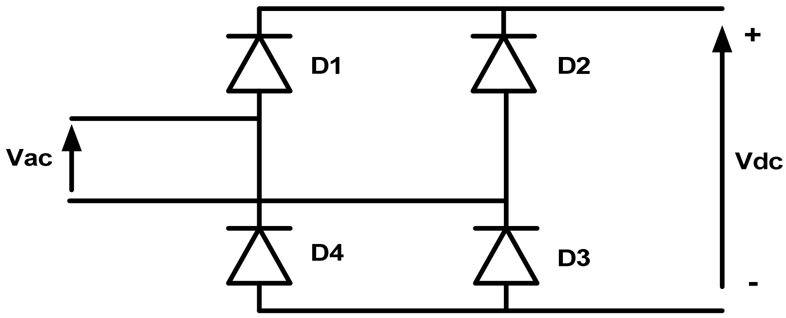


Figure 3 Diode rectifier input bridge

When the input voltage is positive, diodes D1 and D3 are "on" because their anode voltage is higher compared to their cathode voltage. During this phase, D2 and D4 are "off". When the input voltage is negative, it is the contrary, D2 and D4 are "on" and D1 and D3 are "off".

This circuit can be improved by adding of a capacitor in parallel to the rectifier bridge output which gives currents with fewer harmonic. The capacitor can be chosen by using the following expression:

$$C = I_{max}/(\Delta U . f) \qquad \textbf{(4.1)}$$

Where I_{max} is the maximum value of the bridge output current, ΔU the ripple voltage wished and f the frequency.

Other more complicated filters exist and allow having a more continuous voltage than with the capacitor. However, the use of a capacitor is impossible in the first full bride rectifier because of the waveform which has to be sinusoidal (in absolute value).

Figure 4 shows the output voltage waveform of the bridge rectifier.

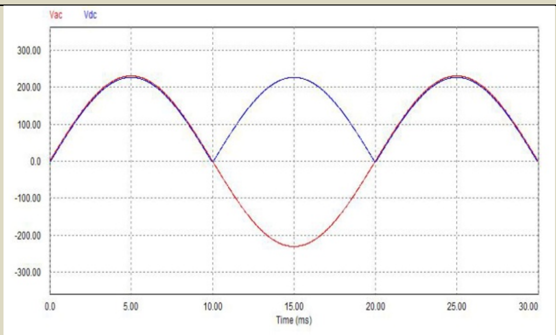
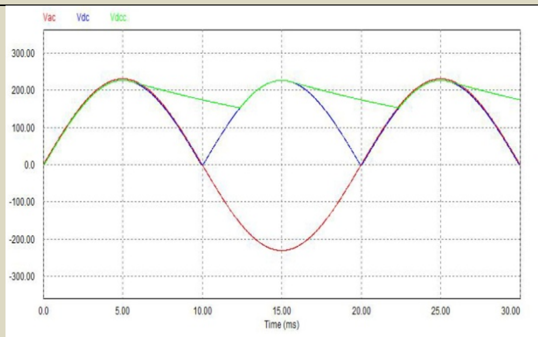
Without capacitor	With capacitor
	
$V_{ac}(t)$ $V_{dc}(t)$	$V_{ac}(t)$ $V_{dc}(t)$ $V_{dc}(t)$with capacitor

Figure 4 Bridge output voltage

DC/DC Stage

The choice of the DC/DC is enough limited in a PFC. The DC/DC converter possibilities are the Boost converter, the Cùk converter or the Flyback converter because an output voltage higher than the input voltage is suitable in a PFC. The high power of the on board charger prevents the use of a Flyback converter. So the boost converter is the most classical structure to carry out the PFC function if an other DC/DC converter to control is required. A boost converter has not a galvanic isolation contrary to the Flyback converter. Figure 5 shows a boost converter.

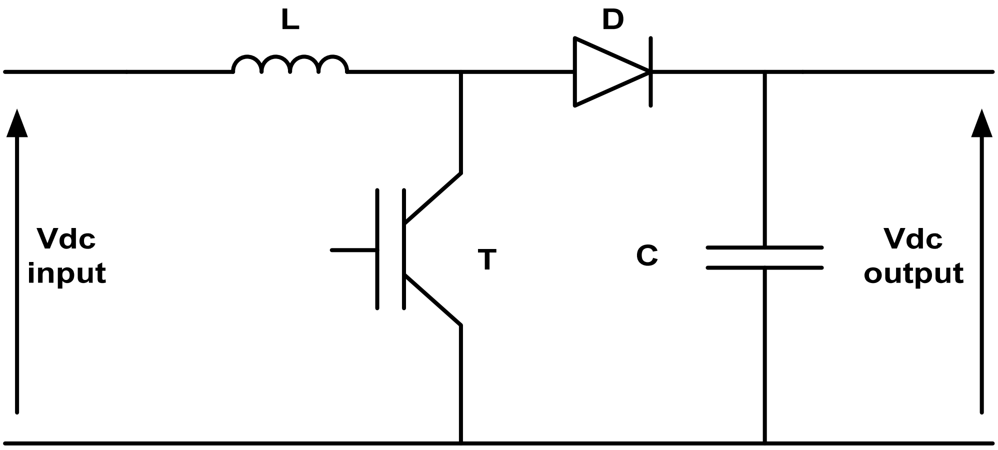


Figure 5 Boost converter

Inductors allows respecting the source interconnection rule. A capacitor allows storing energy and provides energy when the input power in less than the needed output power, in effect, the input power varies between 0 and $V_{max}I_{max}$ and the output power is constant. The operation procedure is detailed below.

First phase: $0<t<\alpha T$ T "on" D "off"

Figure 6 shows the way of the current during the first phase.

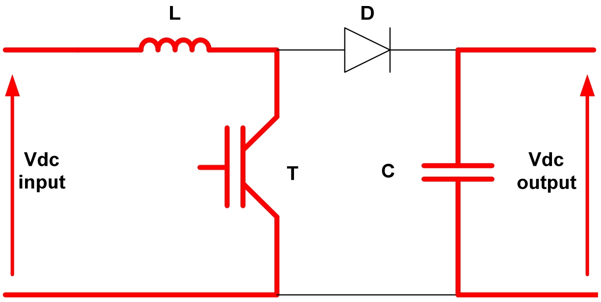


Figure 6 Current during the first phase

- $V_T = 0$ **(4.2)**
- $V_L(t) = L\frac{di_L(t)}{dt} = V_{in}$ **(4.3)**
- $V_D = -V_{out}$ **(4.4)**
- $i_L(t) = i_T(t) = \frac{V_{in}}{L}t + I_{L\,min}$ **(4.5)**
- $i_D(t) = 0$ **(4.6)**
- $i_{out}(t) = -i_c(t) = \frac{V_{out}}{R} = I_{out}$ **(4.7)**

Second phase: $\alpha T<t<T$ T "off" D "on"

Figure 7 shows the way of the current during the second phase.

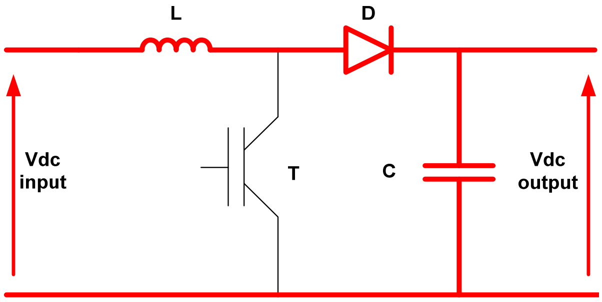


Figure 7 Current during the second phase

- $V_D = 0$ **(4.8)**
- $V_T(t) = V_{out}$ **(4.9)**
- $V_L(t) = V_{in} - V_{out} = L\frac{di_L(t)}{dt}$ **(4.10)**
- $i_L(t) = \frac{V_{in}-V_{out}}{L}(t - \alpha T) + I_{L\,max}$ **(4.11)**
- $i_T(t) = 0$ **(4.12)**
- $i_d(t) = i_l(t) = i_c(t) + I_{out}$ **(4.13)**

The boost converter waveforms can be seen on Figure 8.

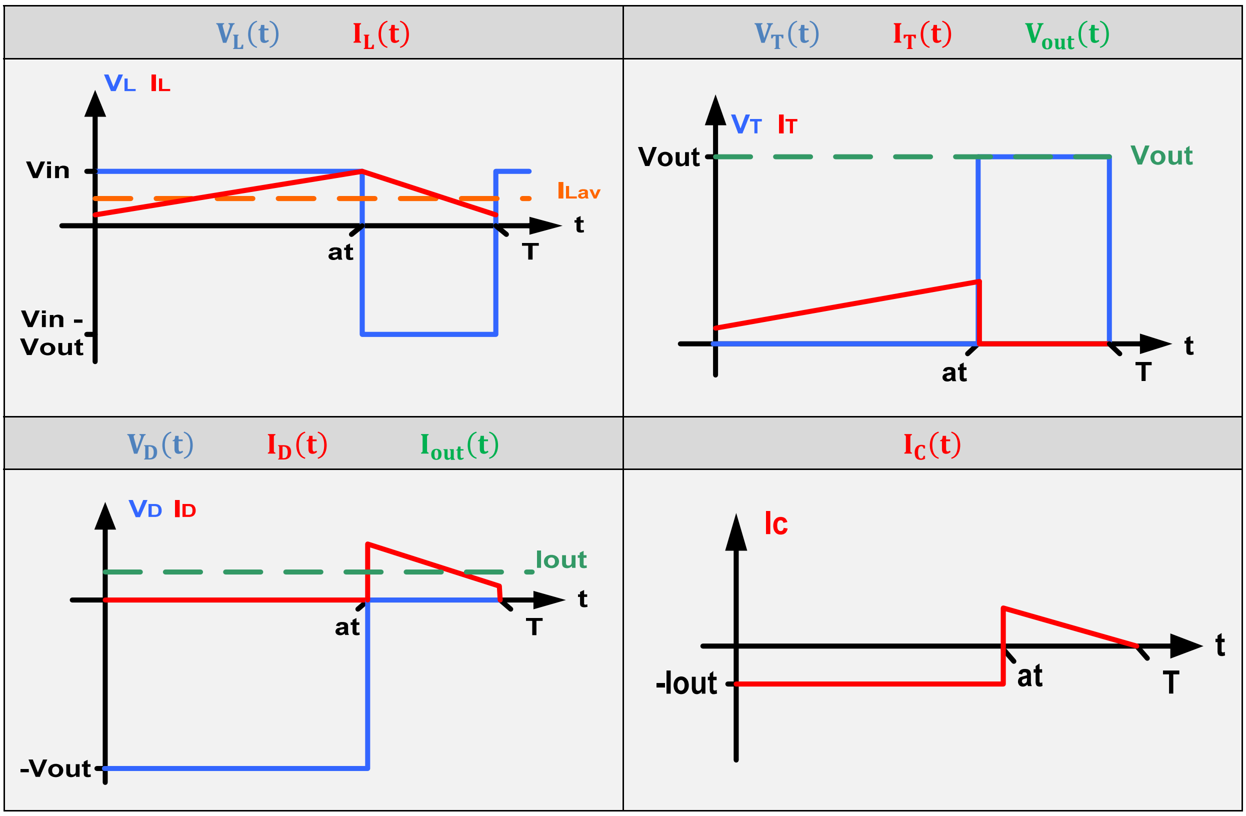


Figure 8 Current and voltage waveform in the boost converter

The relation between the input voltage and the output voltage can be found with the inductor voltage waveform. This relation is:

$$\mathrm{V_{out}} = \frac{1}{(1-\alpha)} \mathrm{V_{in}} \qquad \textbf{(4.14)}$$

Where $\mathrm{V_{out}}$ is the output voltage, $\mathrm{V_{in}}$ the input voltage and α the value of the duty cycle.

If α is equal to 0, the output voltage is equal to the input voltage. For all the other values, the output voltage is higher than the input voltage.

The inductor value is determined by the allowed current variation amplitude and on the switching frequency.

$$L = \frac{V_{out}}{4\, f_{swi\,max}\, \Delta I} \qquad \textbf{(4.15)}$$

Where $\mathrm{V_{out}}$ is the output voltage, $f_{swi\,max}$ the transistor maximum switching frequency and ΔI the ripple current.

The capacitor value is determined by the allowed in load voltage variation amplitude.

$$C = \frac{V_M I_M}{2\, \Delta V_{out}\, \omega\, V_{out}} \qquad \textbf{(4.16)}$$

Where V_M is the maximum input voltage, I_M the maximum input current, ΔV_{out} the output voltage ripple, ω is the pulsation and$V_{out}$the output voltage.

Control stage

The PFC control requires two series of loops. The first one allows controlling the output voltage. The output of this loop will be the reference of the internal loop which allows controlling the form and amplitude of inductor current.

There are different ways to allow controlling the current using a converter. Two main methods are Hysteresis control and Pulse Width Modulation. The hysteresis control is the easiest but has as drawback that the frequency is not fixed. Control with a PWM operating at a fixed frequency but requires more calculus.

In this boost converter, the PWM control is chosen.

Current loop

The current loop allows controlling the waveform and value of the inductor current. First, a study of the hysteresis control is studied and then, the PWM control is presented.

The current loop constant time has to be lower than the one of the voltage loop. So, the cutoff frequency has to have a high value (more than 1000Hz). This value allows considering the current loop like being equal to a unitary gain in relation to the voltage loop.

Hysteresis control

With a hysteresis control (Figure 9**Error! Reference source not found.**), the cutoff frequency is variable. It depends on the inductor value, inductor voltage, and current variation amplitude. So, the choice of the inductor value is very important because of the cutoff frequency value which influences the switching losses.

The inductor value can be chosen using the following expression:

$$L = \frac{V_{out}}{4\, f_{swi\,max} \Delta I} \quad \textbf{(4.17)}$$

Where V_{out} is the output voltage, $f_{swi\,max}$ the transistor maximum switching frequency and ΔI the ripple current.

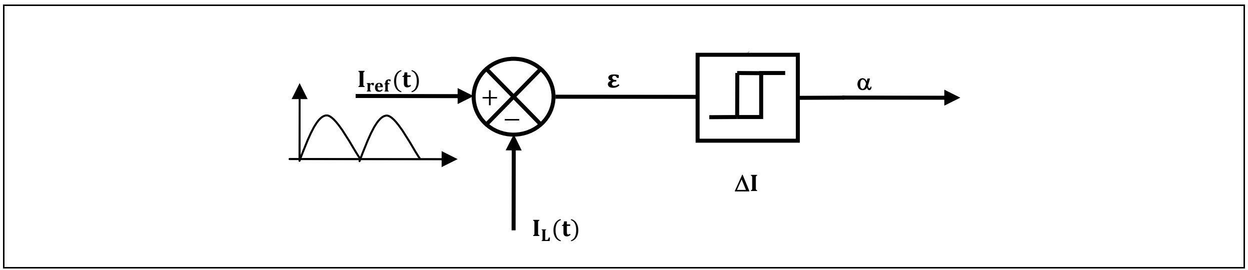


Figure 9 Current control by hysteresis

With a hysteresis control, the real current is kept between two thresholds. When the high threshold is reached, transistor switches off until the low threshold is reached. At this moment, transistor becomes “on”.

PWM control

In contrary to the hysteresis control, a PWM control (Figure 10) has the advantage to work at a cutoff frequency fixed. It is necessary to implement a PI controller.

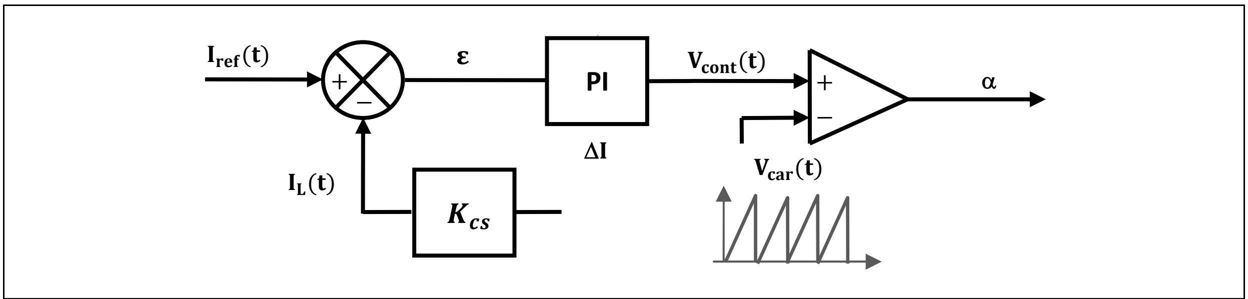


Figure 10 Current control by PWM

PI controller:

A PI controller has the following form:

$$C_{pi}(p) = K\frac{(1+\tau p)}{\tau p} = K_p + \frac{K_i}{p} \qquad \text{where} \begin{cases} K = K_p \\ K_i = K/\tau \end{cases} \tag{4.18}$$

It is the second form which is used for the calculations in this work.

Calculus K_p:

To calculate K_p , the transfer function amplitude has to be equal to 1 at $\omega = \omega_c$.

$$|T_{ol}|_{\omega=\omega_c} = 1 = \frac{K_p K_{cs} V_{out}}{V_{p\,max} L\,\omega_c} \tag{4.19}$$

$$K_p = \frac{V_{p\,max} L\,\omega_c}{K_{cs} V_{out}} \tag{4.20}$$

Where $V_{p\,max}$ is the maximum value of the carrier signal, L the inductor value, ω_c the cutoff pulsation, K_{cs} the current sensor coefficient and V_{out} the output voltage.

Calculus K_i:

To calculate K_i , and in the aim to have a stable system in closed loop, a sufficient phase margin is required at $\omega = \omega_c$. This phase margin determines the system bandwidth.

$$M_{(\Phi)} = -\mathrm{Arctg}(\frac{K_p}{K_i})\ \omega_c \tag{4.21}$$

$$\tau = \frac{\tan(M_\Phi)}{\omega_c} \tag{4.22}$$

$$K_i = \frac{K_p}{\tau} \tag{4.23}$$

Where M_Φ is the margin phase, ω_c the cutoff pulsation.

Voltage loop

The voltage controller (Figure 11) response process has to be slow enough in comparison with the 100Hz with the aim to keep the sinusoidal inductor current waveform. The voltage loop cutoff frequency has to be in the order of 10 Hz.

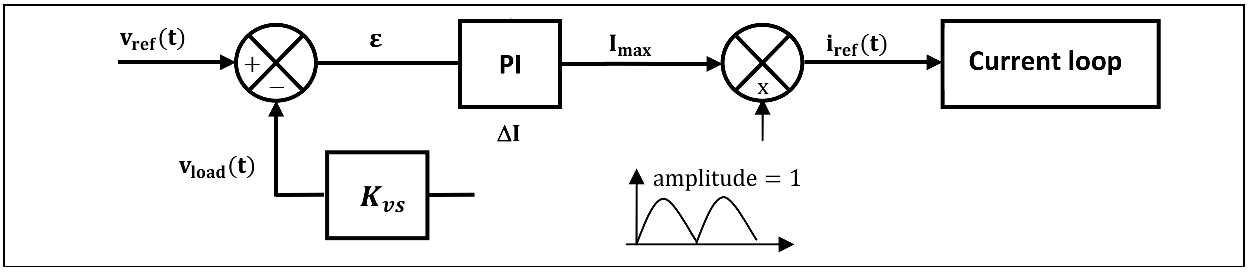


Figure 11 Voltage control

PI controller:

A PI controller has the following form:

$$C_{pi}(p) = K\frac{(1+\tau_v p)}{\tau_v p} = K_p + \frac{K_i}{p} \qquad \text{with} \begin{cases} K = K_p \\ K_i = K/\tau_v \end{cases} \tag{4.24}$$

Calculus τ:

To calculate τ, the controller time constant has to be equal to the one of the process. That is dominant pole compensation.

$$\tau_v = \frac{RC}{2} \tag{4.25}$$

Where R is the resistor value and C the capacitor value.

Calculus K:

To calculate K , it is necessary to impose the closed loop transfer function cutoff frequency (f_c about few Hertz).

$$\omega_c = \frac{K_{vs}\, R\, V_{in\,max}\, K_v}{4\, V_{out}\,.\tau_v} \tag{4.26}$$

$$K_v = \frac{4\, V_{out}\, \tau_v\, \omega_c}{K_{vs}\, R\, V_{in\,max}} \tag{4.27}$$

Where V_{out} is the output voltage, τ_v the system time constant, ω_c the cutoff pulsation, K_{vs} the voltage sensor coefficient, R the resistor value and $V_{in\ max}$ the maximum input voltage.

Pulse transformer

A pulse transformer is used to give pulses to the gate transistor. This component is placed between the FPGA and transistor. It often has a transformer ratio of 1 or 2.

For a good operation of the pulse generator, the iron should not be saturated. To avoid the iron saturation, the transformer value of the product of voltage by time (VT) is important. This constant is given by the manufacturer and, mainly, varies between a few Vµs and 200Vµs. It also is necessary to accomplish a complete demagnetization phase between each active phase. A diode in parallel at the transformer primary allows the demagnetization. A zener diode, in series with this diode, makes the demagnetization phase quicker. The pulse transformer can be controlled by µC or DSP or FPGA to send the pulse at the angle or time required.

In a lot of power electronic applications, the pulse transformers are used for the control of the triggering of thyristors, triacs or transistors. It has many advantages: it can work at high frequencies, simple usefulness, and it can supply a high current.

PFC Improvement

It is possible to improve this circuit by creating a two phase PFC. This PFC structure is shown in Figure 12 This kind of PFC allows for a reduction of the input and output current ripple, generate lower distortion sinusoidal input line current and consequently, make the EMI filter lighter and less expensive. The difficulty lies in the fact that the two transistors have to be control with delay between them. In effect, the second transistor has to be delayed of 180° toward the first transistor. This kind of control can be realized using a FPGA, but it also exist circuits which realized this control, i.e. measuring the both inductor currents. It also exist four phase PFC realized on the same principle. In this case, the transistor control is delayed of 90°. These circuits can also improve the efficiency of the on-board charger.

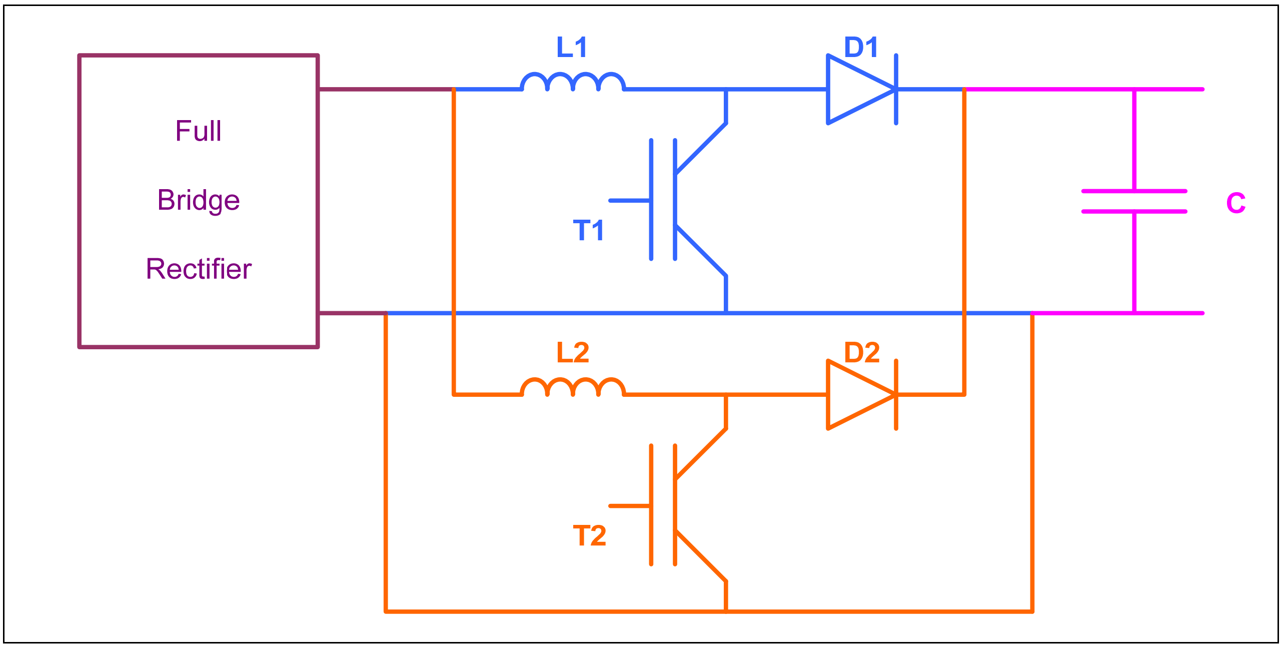


Figure 12 Two phases PFC

B) DC/DC Converter

The choice of the DC/DC converters depends on several factors. Mainly, it depends on the power which has to be transmitted and the knowledge if a step down or step up converter is required. It is also necessary to know if isolation is needed. In the OBC the converter has to be a step down stage, in effect, the input voltage is higher than the output voltage. A galvanic isolation is required, so a transformer has to be used. All these information make the following choice for the structure which can be viewed on Figure 13.

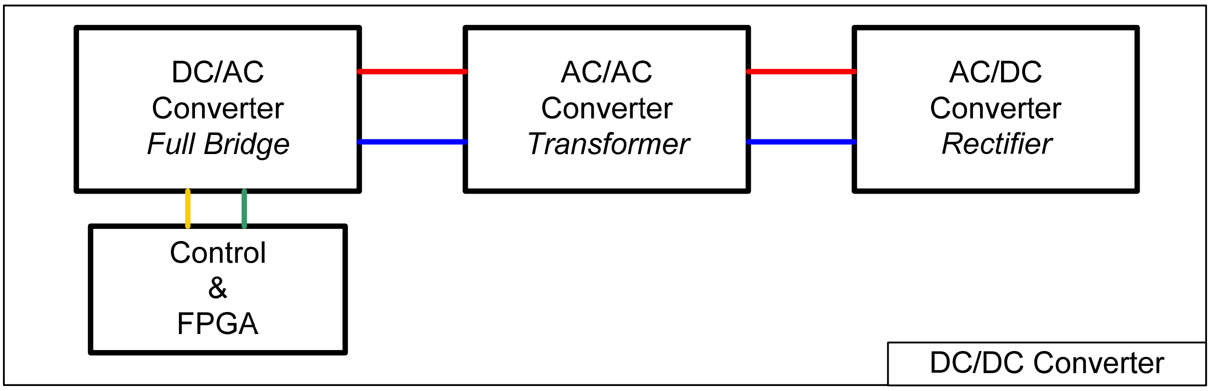


Figure 13 Structure of the full bridge converter

DC/AC Stage

This stage is an inverter device. It converts a continuous signal in an alternative signal. The choice of a full bridge converter (Figure 14) is guided by the power to transmit. With a full bridge inverter, the maximum output voltage is twice that a half bridge inverter. For the same power, output and switch current are half one of those for a half bride inverter. The transistors are smaller in the full bridge inverter. With very high power, a full bridge inverter required less parallel devices than a half bridge inverter.

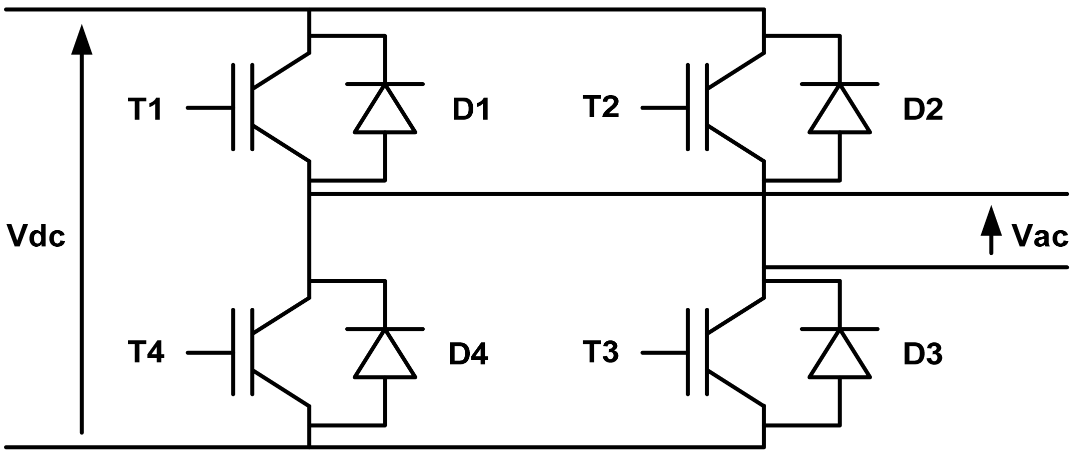


Figure 14 Full bridge inverter

The full bridge converter working is simple. Each transistor is "on" during half period and "off" during the other half period. Switch T_1 can't be "on" in same time that T_4. There are two different ways for the full bridge converter control:

- Simultaneous: T_1"on" = T_2"off" = T_3"on" = T_4"off"
- Delay: T_1"on" = T_4"off" and T_2"on" = T_3"off"

The difference between these two controls can be seeing below in the waveforms (Figure 15) where the focus is on the output voltage according to control method.

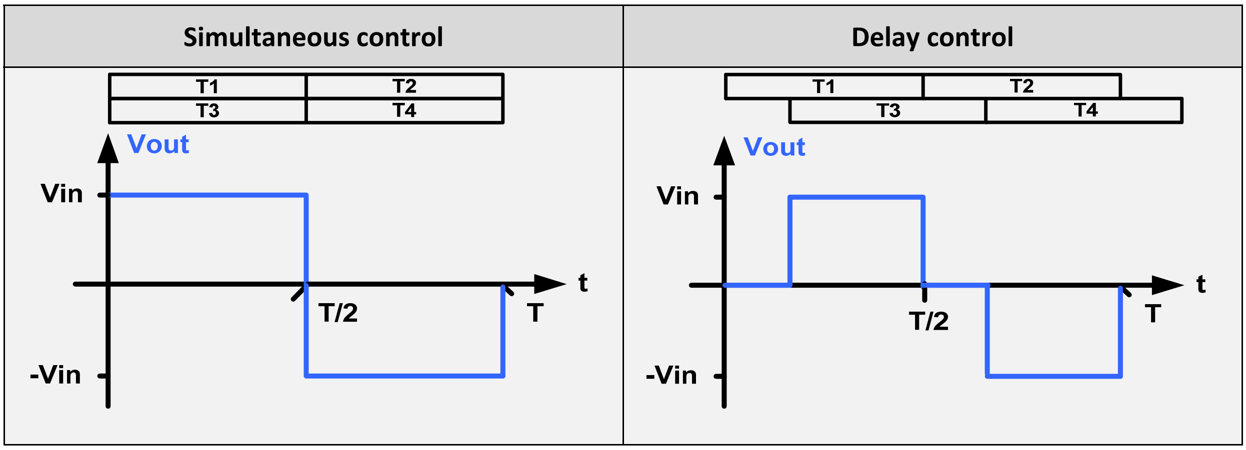


Figure 15 Output voltage according to the control method

Components

Transistor

A transistor is a solid-state semiconductor device used as a switch in the power electronic field, mainly in the power electronic circuit. There are several different kinds of transistor.

The choice of the transistor is important, if a wrong type is used and the on board charger can't work correctly and have higher losses. The choice between BJT (bipolar junction transistor), MOFSET (metal oxide semiconductor field effect transistor) and IGBT (insulated gate bipolar transistor) is accordingly an important issue.

The BJT is nowadays not used in power electronics power circuits; this kind of component is most useful in current amplifier. The choice between a MOSFET and an IGBT is more difficult, these both components can be a good choice in this supply. For a switching frequency of 20 kHz, the IGBT is probably the best choice but the MOSFET can work here too. For the voltage of 400V, IGBT is also probably better than the MOFSET but MOSFET can also be found this voltage. So, the best transistor is probably the IGBT in this supply.

IBGT Transistor:

This transistor has the advantages of the BJTs and the MOSFETs. It has the switching simplicity of the FET transistors (Field Effect Transistor) and the low conduction losses of the bipolar transistor. It presents the field effect transistor characteristics with regard to the control of the device and bipolar transistor characteristics for the output. The IGBT transistor has a higher switch off time than MOSFET, for this reason the switching frequency is lower than for the MOFSET (1). The Figure 16 shows the two different symbols of an IGBT.

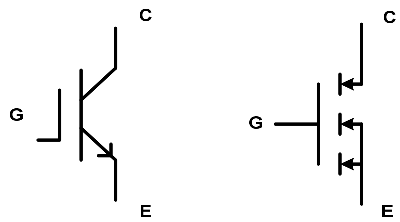


Figure 16 IGBT symbols

This IGBT structure (Figure 17) is quite similar to the MOFSET. The main difference is the presence of a p^+layer which forms the drain of the IGBT. This layer creates a PN junction that means that the IGBT cannot conduct in the reverse direction contrary to the MOSFET. The p^+ layer allows for a reduction of the drop voltage during the "on" state. The n^+layer plays the role of buffer, making the switching frequency higher. There is also a PNP junction between source and drain like in a bipolar transistor and a n^+layer between gate and source like in the MOSFET transistor (2).

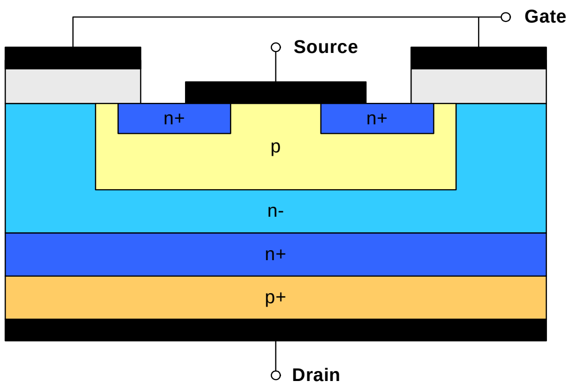


Figure 17 IGBT structure

Their switching method is simple. The IBGT transistor works like FET transistor in regard with its control so it is switched by the gate voltage (voltage between Gate and Emitter). The IGBT turns on when the collector-emitter voltage is positive and a positive signal is applied at the gate input. It turns off when the collector-emitter voltage is positive and no signal is applied at the gate input.

MOSFET Transistor:

MOSFETs have replaced BJTs in many applications especially where high switching frequency is needed. At high voltage (>600 - 1000 V), IGBTs are preferred to MOSFETs, but at high frequency (>20 - 100 kHz), MOSFETs are the only power transistors which can be used. Figure 18 shows the MOSFET transistor symbol.

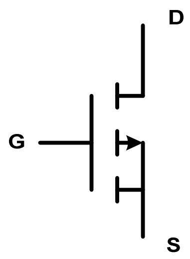


Figure 18 MOSFET symbol

The MOSFET structure, shown in Figure 19 consists on a vertically layer structure of alternative p-type and n-type doping. There is a NPN junction between source and drain. The MOSFET structure allows it to conduct in the reverse direction unlike the IGBTs. The gate is isolated from the source and drain by a dielectric isolation layer.

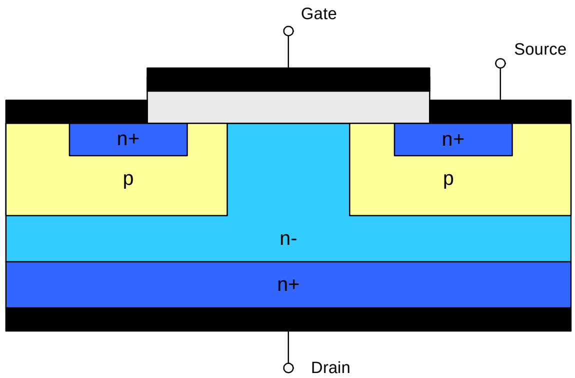
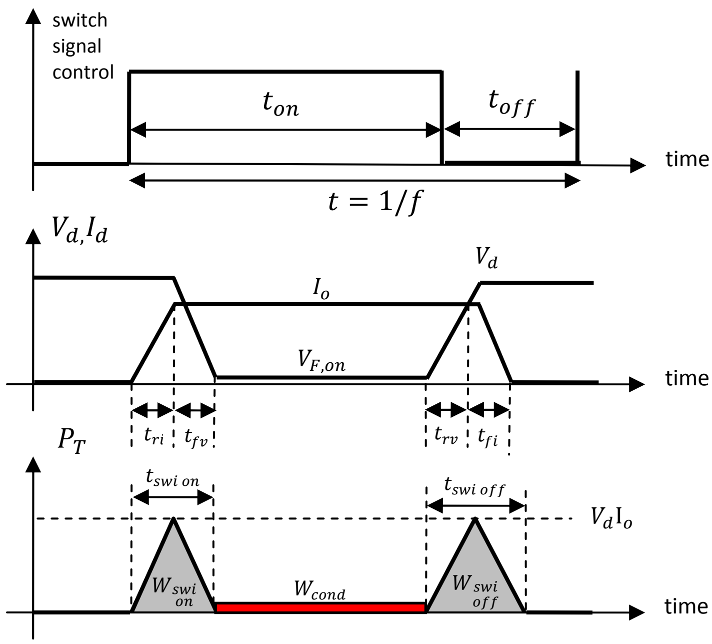


Figure 19 MOSFET structure

Like the IGBT, the MOSFET switches on when drain-source voltage is positive and when a positive signal is applied to the gate. It switches off when drain-source voltage is positive and no signal is applied at the gate.

Figure 20 shows the switching waveforms and the switching losses for the transistor.



$$W_{swi\,on} = \frac{1}{2} V_d I_o t_{swi\,on} \tag{4.28}$$

$$W_{swi\,off} = \frac{1}{2} V_d I_o t_{swi\,off} \tag{4.29}$$

$$W_{cond} = V_{F,on} I_o t_{on} \tag{4.30}$$

Figure 20 Current and voltage in the transistor

Diode

It is necessary to use a fast diode in the full bridge converter due to the switching frequency. These diodes are designed to be used in very high frequency circuit where a very low reverse recovery time t_{rr} is required. At power levels of several hundred of volts and several hundreds of amperes, such diodes have a reverse recovery time less than a few microseconds (2). The fast diode are a supplementary advantage, their switching losses are lower due to their very short reverse recovery time. Indeed, the switching losses are due to the reverse recovery current I_{rrm} and the reverse recovery time. Figure 22 shows the waveform during the reverse recovery phenomenon. The conduction losses are the cause of the forward voltage drop V_F during the on state. This voltage can be seen on the current/voltage characteristic (Figure 21).

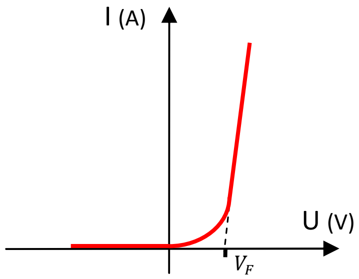


Figure 21 Current/voltage diode charasteristic

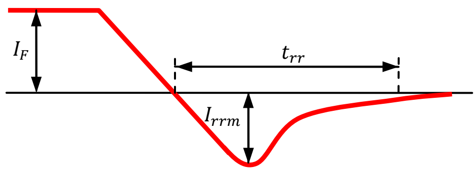


Figure 22 reverse recovery time phenomenon

Snubbers

The aim of a snubber circuit is to reduce the electrical stresses placed on the semiconductor during turn on and/or turn off switching of the transistor. Snubbers limit the voltage applied to devices during turn off phase, limit current during turn on phase, and limit the rise of dv/dt and di/dt during turn on and turn off phases (2). For transistors, there are 2 different snubber circuits:

- Polarized R-C snubbers used to shape the turn off portion of the switching trajectory of transistor, to clamp voltages applied to the transistor to safe level and to limit dv/dt during turn off phase.
- Polarized L-R snubbers are used to shape the turn on portion of the switching trajectory of transistor and to limit di/dt during turn on phase.

In the full bridge inverter, the focus is on the turn off snubber. Figure 23 shows a capacitor snubber and Figure 24 a capacitor and resistor snubber. This one can reduce the turn off losses for the IGBT modules (3). Its goal is to provide a low voltage across the transistor while the current turns off.

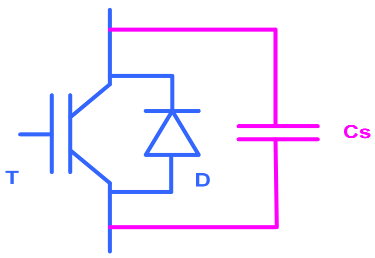


Figure 23 Capacitor snubber

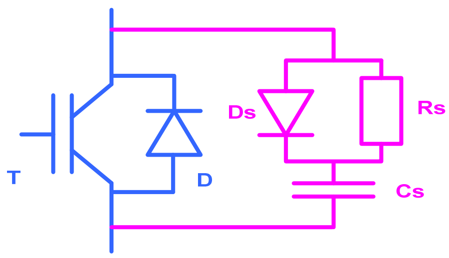


Figure 24 Resistor, capacitor and diode snubber

The first solution for this snubber is a capacitor connected across the switch (Figure 23). The capacitor is discharged into the transistor and has to be discharged before the transistor is switched on. This solution is the simplest but doesn't take the turn on phase into account. In addition, the presence of the snubber capacitor causes an increase of the current during the turn on phase. A second solution consists of a RCD snubber (Figure 24) where the resistor limits the current peak during the turn on phase.

This solution has several beneficial effects:

- The capacitor energy is dissipated in the resistor (resistor is easier to refresh).
- No additional energy in the transistor due to the turn off snubber.
- The turn on peak current is not increased due to the turn off snubber.

The drawback is that there are losses in the resistor when the capacitor is completely discharged.

The kind of snubber also depends on the converter and the control method and a capacitor snubber is probably the best choice for our system.

Control stage

The control stage of this converter allows for the controlling of the battery voltage and current. A battery requires a first phase at constant current and a second phase at constant voltage. So, it is necessary to maintain these two quantities at selected values.

The easiest way to control the full bridge converter is the duty cycle control. With this method the output voltage is proportional to the duty cycle value. This method gives high switch losses due to the switches which are turned off at full load voltage. There are hard-switching conditions at both turn-on and turn-off for all transistors (3). A PWM with bipolar voltage switching pattern is used in this control.

A second way to control this converter, more difficult, is the phase shift control. With this method, the two switching legs are controlled separately. This method gives less switch losses if there are snubber circuits across the switches (3). A PWM with unipolar voltage switching is used for phase shift control.

A comparison between these two kinds of PWM allows for the selection of the best one according to the losses in the semiconductors.

Current loop and PWM

The current loop enables for the control of the waveform and value of the inductor current. The unipolar voltage switching control (phase shift control) is chosen for the first control of the full bridge converter. With this PWM method, the switches in the two legs are not switching simultaneously. The first leg (S1, S4 Figure 14) is switched by comparing $V_{control}$ with $V_{triangle}$ and the second leg (S2, S3 Figure 14) is switched by comparing $-V_{control}$ with $V_{triangle}$.

- $V_{control} > V_{triangle}$: T_1 "on"
- $V_{control} < V_{triangle}$: T_4 "on"
- $-V_{control} > V_{triangle}$: T_2 "on"
- $-V_{control} < V_{triangle}$: T_3 "on"

Figure 25 shows the current loop with phase shift control.

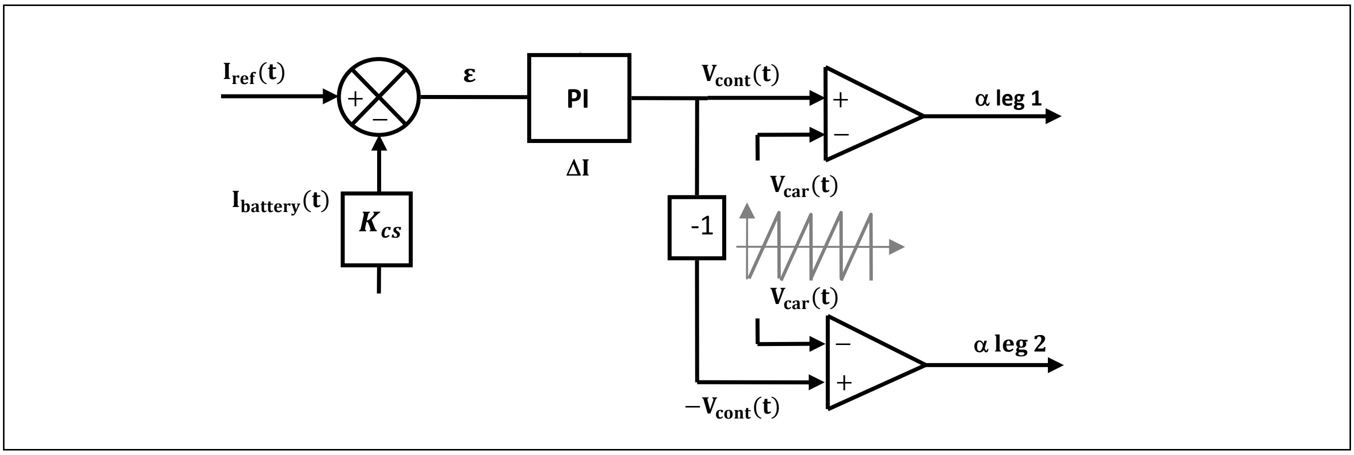


Figure 25 Current loop with phase shift control

The second control method is the bipolar voltage control (duty cycle control). With this method, the diagonally opposite switches (T_1 T_3 and T_2 T_4) are switched as pairs. The output of the leg 2 is negative of the leg 1 output.

For *t* between 0 and T/2:

- $V_{control} > V_{triangle}$: T_1 T_3 "on"
- $V_{control} < V_{triangle}$: T_1 T_3 "off"

For *t* between T/2 and T:

- $V_{control} > V_{triangle}$: T_2 T_4 "on"
- $V_{control} < V_{triangle}$: T_2 T_4 "off"

Figure 26 shows the current loop with duty cycle control

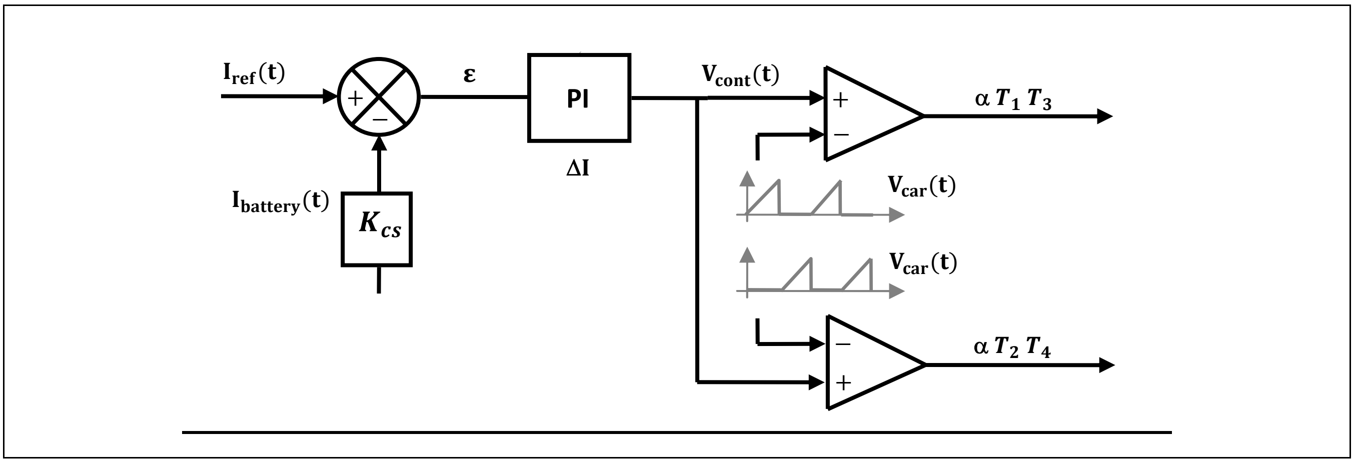


Figure 26 Current loop with duty cycle control

AC/AC Stage

A transformer consists of two or more coils which are magnetically coupled. In this converter, transistor are switching at a high frequency (about 20 kHz) so the inverter output signal has a high frequency. The transformer universal EMF equation says:

$$E = K\ N\ f\ a\ B \qquad \textbf{(4.31)}$$

Where *E* is the winding voltage, *f* is the frequency, *N* the number of turns, *a* the core cross-sectional area and *B* the peak magnetic flux density, K is 4.44 for sinusoid and 4 for rectangular wave (4).

As a consequence of this relation, if the frequency is high, the core section is low, and vice versa. Operation at high frequency causes other problem.

The Eddy current losses are proportional to square frequency and in inversely proportional to transformer material resistivity. To limit the Eddy current losses, the magnetic circuit of the high frequency transformer is realized using an insulating ferromagnetic material. Since the DC/DC converter is a full bridge, the high frequency isolation transformer has to have a bidirectional core excitation. The transformer also depends on the kind of the AC/DC converter which is used. It will be a middle point transformer or a “classical” transformer. Both have to be able to work at high frequency. But in this DC/DC converter, a 2 winding coils transformer is required (reasons are discussed in the next part AC/DC Stage).

AC/DC Stage

For this stage, there are 2 ways to achieve this converter. The first one is to use a middle point transformer with a diode half bridge and the second one is a transformer with a diode full bridge behind. The second method is chosen in this work. In addition, this kind of converter has a lower diode voltage than the half bridge. Another point can be added, this type of converter allows for a better transformer winding current flowing.

This topology has already been seen in the PFC part (see: Full bridge rectifier)

2) Bidirectional on board charger

This converter makes energy transfer possible two directions, from the grid to the battery and from the battery to the grid. So, this converter can be also used as a supply or can give power to the grid when it needs extra power.

Structure

Figure 27 shows a general diagram of the on board charger with the different converters, and also with the load (battery) and the grid.

Figure 28 shows the power part of the on board charger different converter

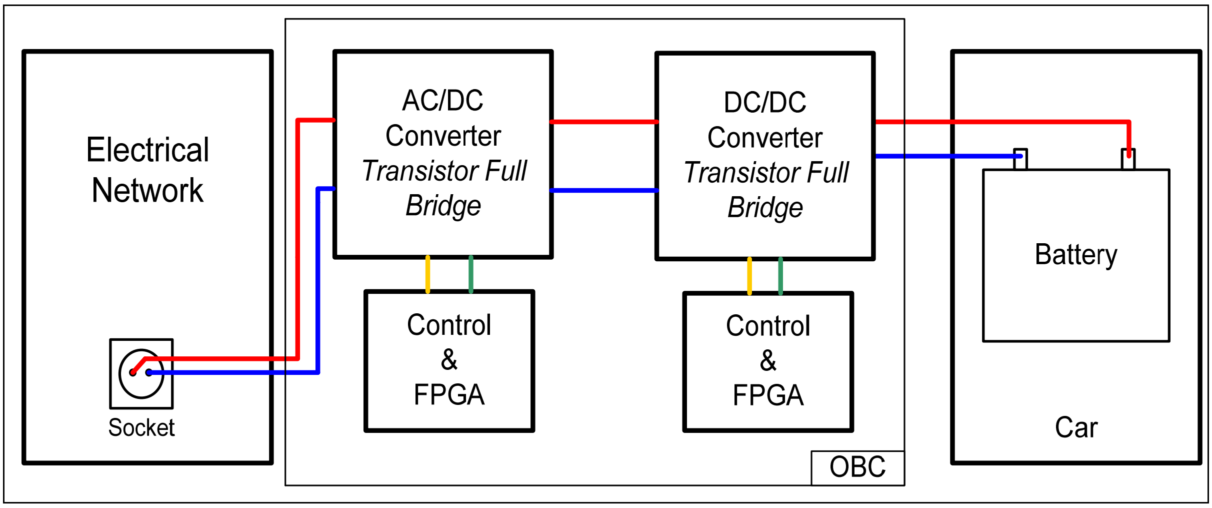


Figure 27 General diagram of the bidirectional on board charger

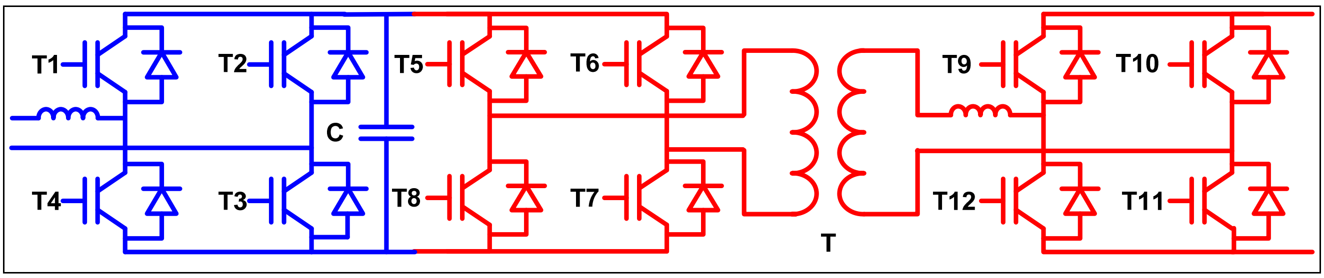


Figure 28 Power part of the bidirectional on board charger

3) Battery

A theoretical study of the load is necessary to design the on-board charger. In addition, voltage and capacity have an influence on the kind of DC/DC converter which is used in the PFC. It's also needs to know how the battery has to be charged and, for the simulation, how batteries can be represented.

Characteristics

The battery voltage varies according to the kind of the plug-in hybrid vehicle. For a car, this voltage could for example be between 300V and 400V. The peak power is in the region of 50kW, and the energy density could be between 5 and 20kWh. For information, an electric car of ordinary size needs about 2kWh to cover 10km. The batteries can have about 1000 cycles of charge. These figures are rough values, and depending on the battery technology, the usage of the battery (small car, truck…) and other parameters.

Battery technology

Nowadays, the batteries that are most used in the hybrid vehicles are the NiMH (nickel-metal hydride) technology. An advantage of this technology is that NiMH batteries are not very sensitive to the memory effect, moreover they have a good energy to weight ratio and are safe. The drawbacks are that overload has a negative effect, and that it is difficult to detect the end loading. However, in a near future, this technology will be surpassed by the Lithium-ion technology. This technology has advantages like not to be effect by memory effect, very good energy to weight and energy to size ratios. The drawbacks are these batteries are less safe than NiMH, they are a higher cost, and a lower load and discharge current. For the moment, this technology is mainly used for small batteries supplying laptops and mobile phones for example.

Charge

The charge of a NiMH battery can be divided into two parts. The first part, the battery is discharged; it needs a constant current to be charged. During this phase, the battery voltage is increasing. This phase stops when the battery reaches its rated voltage. The second phase consists in charge the battery with a constant voltage. The voltage doesn't increase but on the other hand the energy density is increasing. This phase stops when the battery is fully loaded; it doesn't take any current more. The best method to detect the end load is the ΔV method. In effect, when a NiMH battery is full, if the charger continues to supply it, the polarity voltage is reversing (5). If we can detect this inversion, we can know when the battery is full. A charger can be safer by the use of a resettable fuse which opens if the current or temperature is too high.

3) Simulation

With the aim of understanding the working principle of the soft starter, the carrying out of a simulation is necessary. This achievement is realized with Matlab Simulink©. Simulink is an environment for simulation, modelling and calculus in a lot of different dynamical domain. The interface is a graphical block diagramming and is customizable with block libraries. It offers integration with the Matlab environment and is able to drive Matlab or use script from Matlab.

1) Power Factor Corrector

The first simulation realized was the power factor corrector. The power part (Figure 29) was made with an IGBT. The control part (Figure 30) of this converter is realized in order to that the power part realizes the PFC function, which means a sinusoidal input current and an input current in phase with the grid voltage.

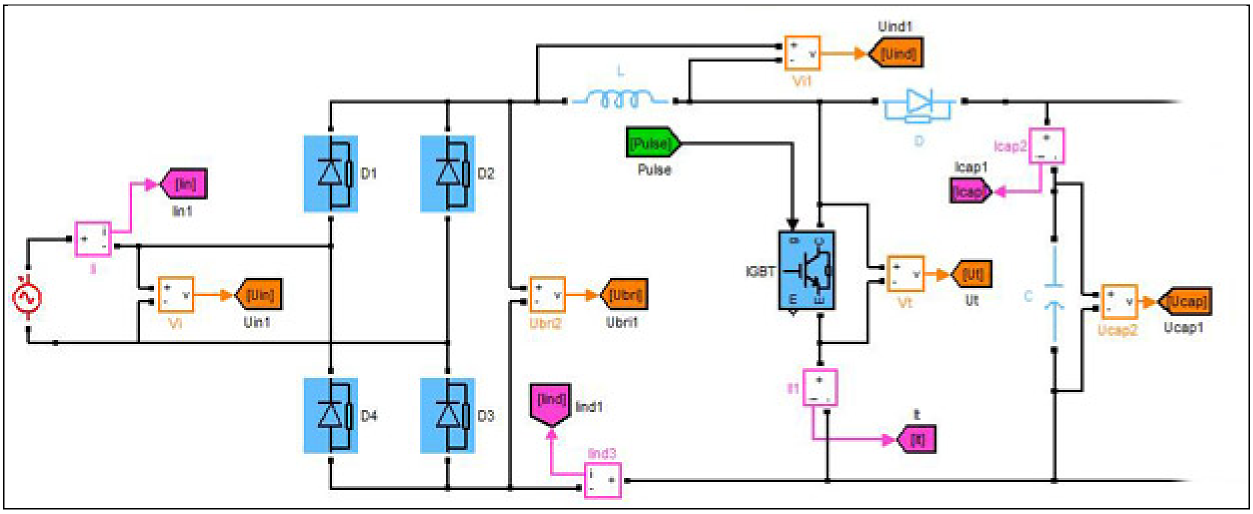


Figure 29 Power part of the PFC

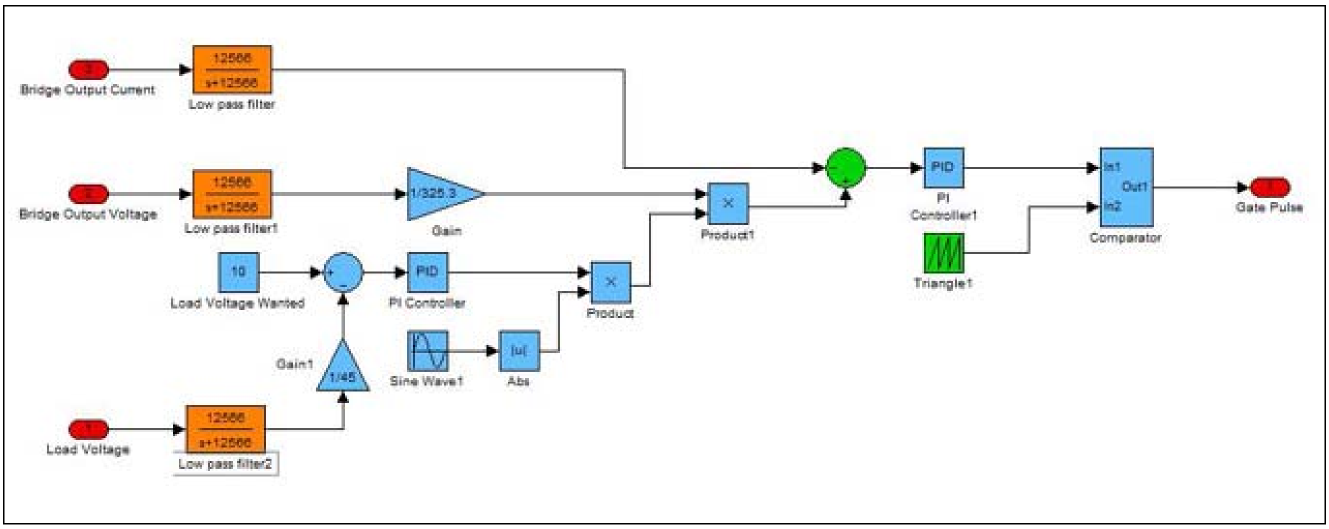


Figure 30 Control part of the PFC

The input current (Figure 31) is a very interesting signal.

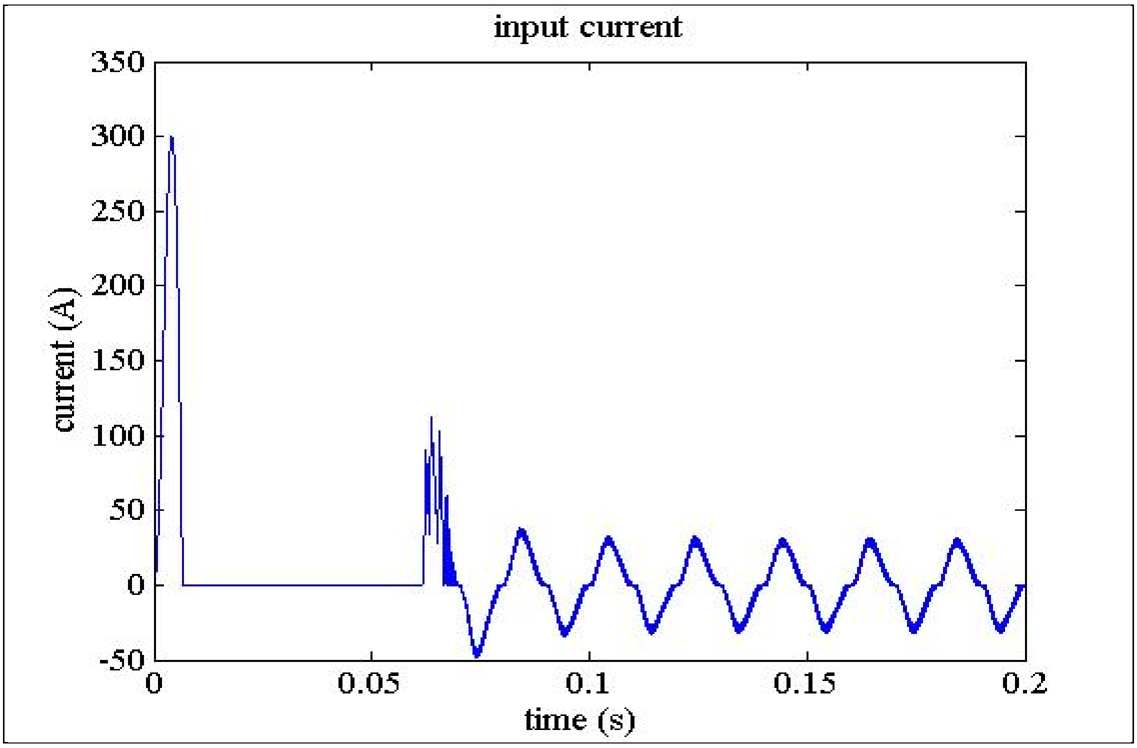


Figure 31 Input current

The input current is almost sinusoidal during the steady state. A high in rush current is present during 7 milliseconds. This current is caused by the capacitor load. In effect, to control the inductor current in the PFC, the capacitor voltage has to be higher than the output voltage of the rectifier bridge. That means that the inductor current increase until the capacitor voltage reached the output voltage of the rectifier bridge. The problem of the in rush current can be solved by several methods:

- A resistor in series between the grid and the input rectifier bridge (Figure 32). This resistor limits the in rush current during the starting phase. When the input current reaches its steady state, the resistor is shunted by a switch in order to eliminate the losses in this one.
- A resistor in series with the DC link capacitor and a switch (FET) in parallel with this resistor (Figure 33). When the input current reaches the steady state, the FET shunts the resistor to eliminate the losses in the resistor.

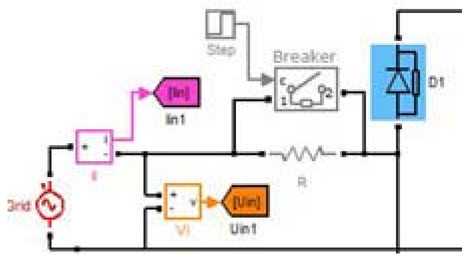


Figure 32 In rush current solver

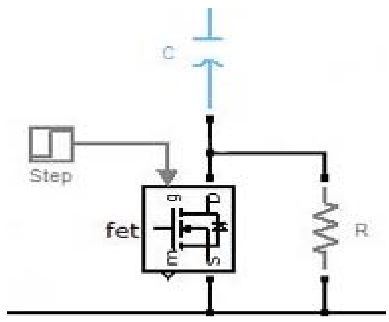


Figure 33 In rush current solver

The Figure 34 shows the input current of the on-board charger and the grid voltage. The input current is in phase with the grid voltage, which means that only an active power is used.

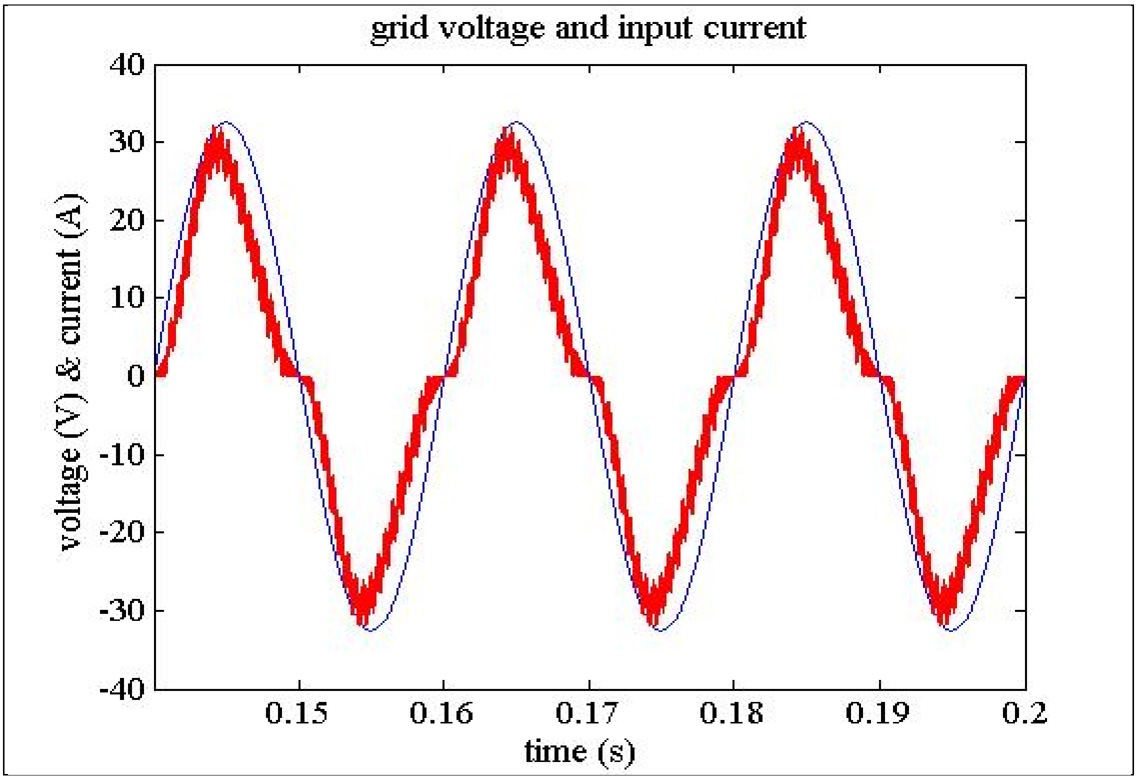


Figure 34 Input current and grid voltage

The current waveform is not perfect, the add of an EMI filter is required. The Figure 35 shows the comparison between the input current without EMI filter and the input current with EMI filter.

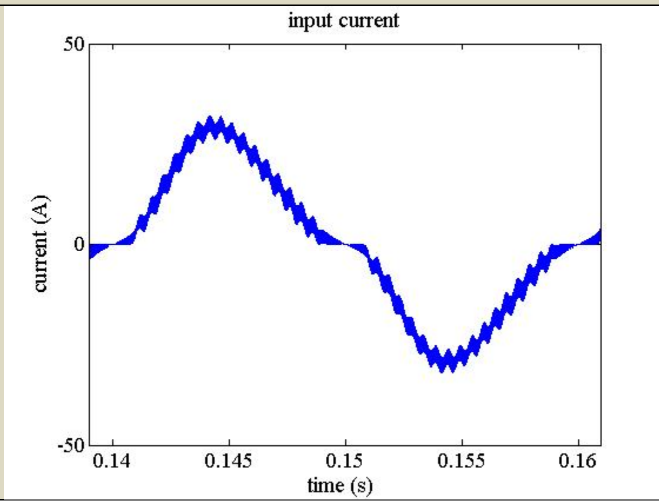
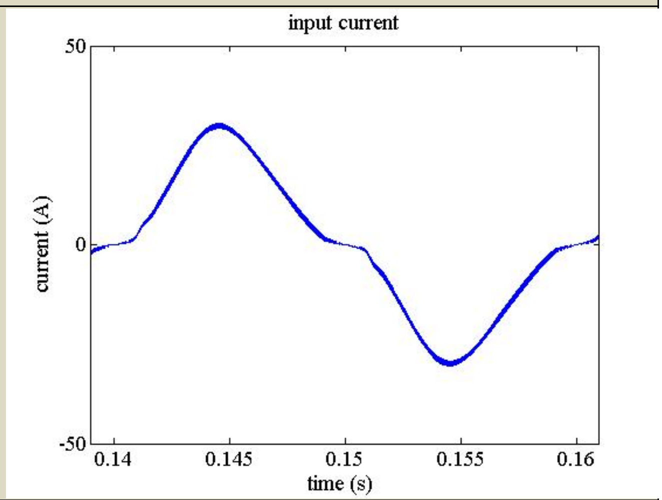
Input current without EMI filter	Input current with EMI filter
	

Figure 35 Role of the EMI filter

The role of the EMI filter is shows on the precedent simulations. When the on-board charger is equipped of an EMI filter, the ripple current are reduced. There are less harmonics and the on board charger is less polluting for the grid.

The waveforms inductor current (Figure 36) is important in a power factor corrector because this current is the reference current in the control of the PFC.

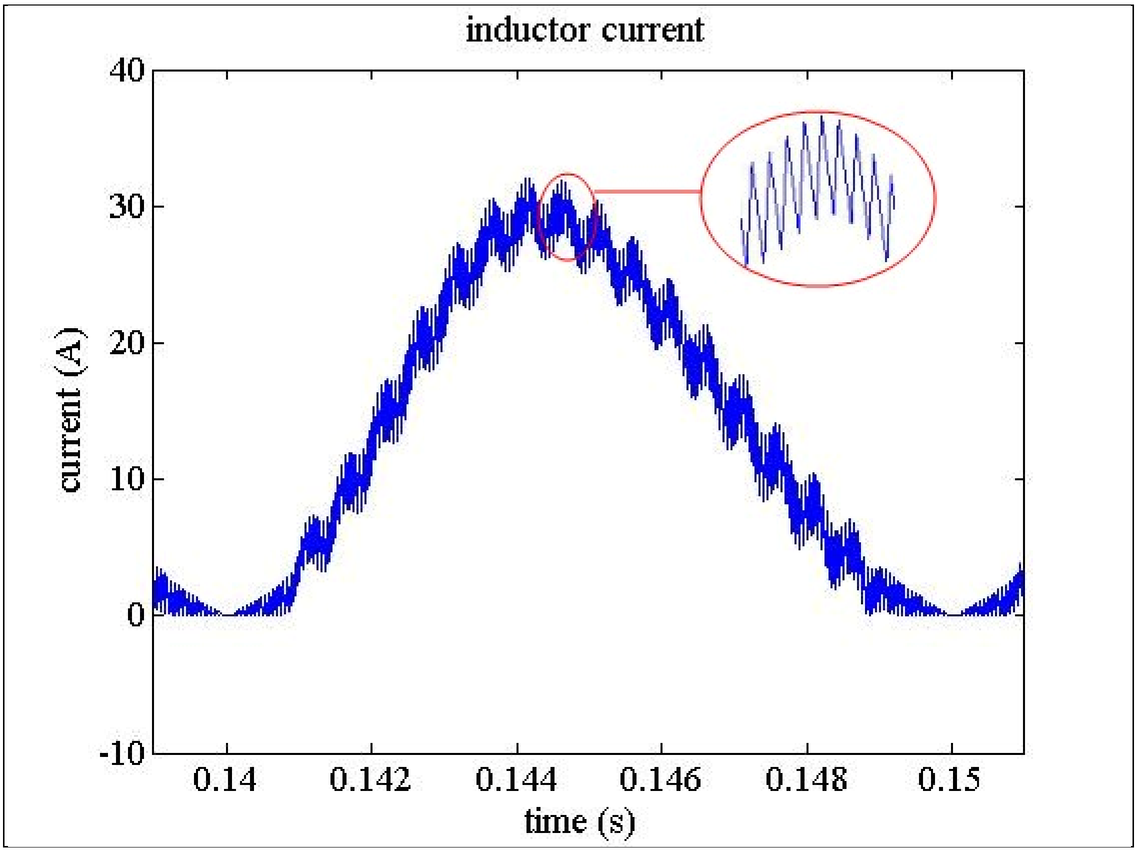


Figure 36 Inductor current

The current ripple depends on the value of the inductor. The inductor value is chosen in order to have a ripple current of 20%. The switching frequency can be seen in the ripple of the inductor current. The waves on this current are due to the fact that the DC link voltage has ripple.

The output voltage of the power factor corrector (DC link voltage capacitor)(Figure 37) is also an important signal. This voltage is the input voltage of the full bridge converter.

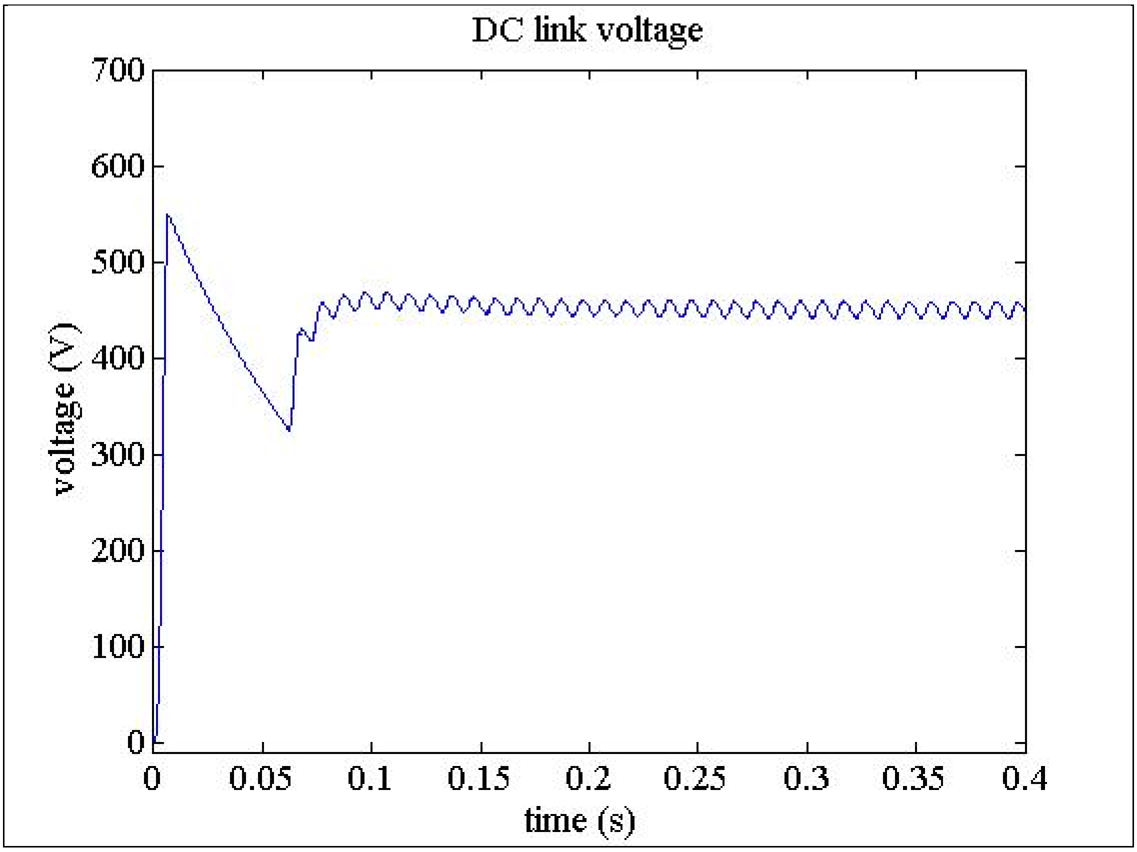


Figure 37 DC link voltage

The value required for the DC link voltage is 450V. This voltage is reached after 200 milliseconds. The output voltage can't be controlled during the load of the capacitor. This phase is occurring during the 70 first milliseconds. To avoid any problem, the full bridge converter will be switch on only after 200 milliseconds; this time is needed to reach the steady state of the power factor corrector.

2) Full bridge converter

The second simulation realized was the full bridge converter. The power part (Figure 38) was also made with IGBTs, snubbers were designed. The control part (Figure 39) realized is the phase shift control method. This one is realized by a unipolar voltage switching control. This converter controls the value of the current in the battery in order to load this one with a constant current.

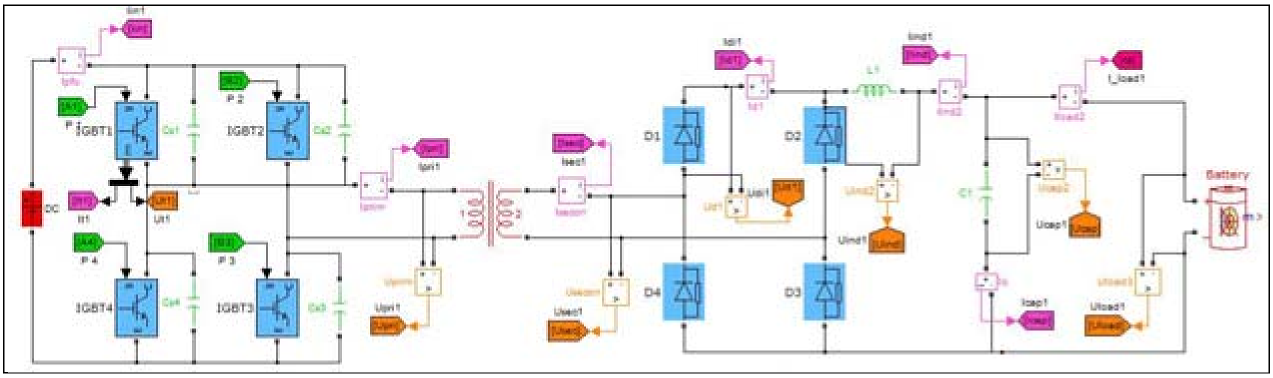


Figure 38 Power part of the full bridge converter

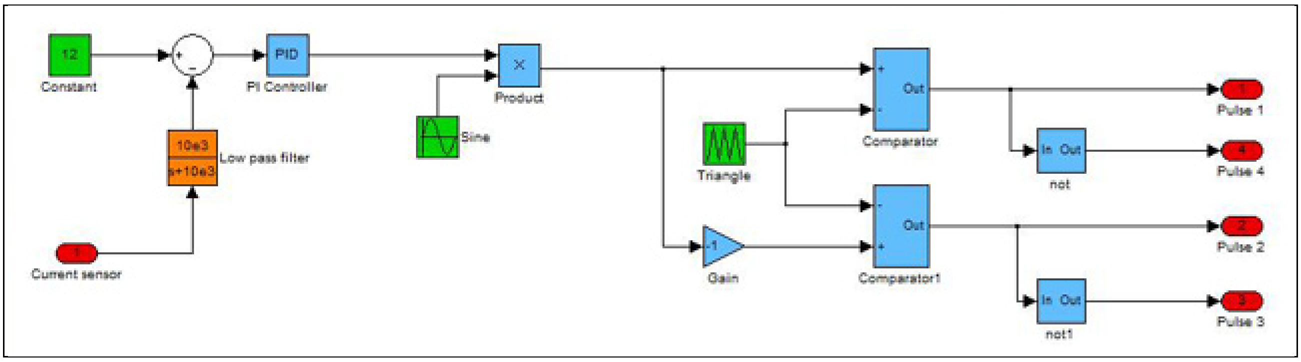
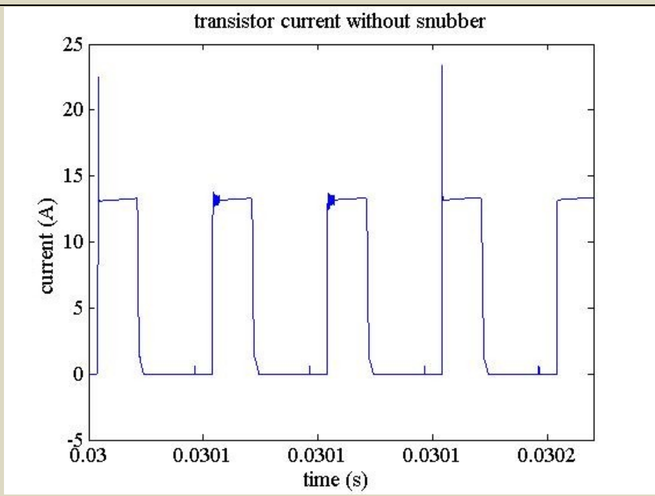


Figure 39 Control part of the full bridge converter

In this converter, the interesting signals are the signals in the transistors, in the transformer, and in the output filter and the battery. These signals can be viewed below.

The Figure 40 shows the current in the transistor T1, with or without snubber. A pure capacitor can't be the only component of this snubber because simulink models the transistor as a current source. The add of a resistor is needed. The Figure 41 shows the voltage in the transistor T1 in both cases, with or without snubber

Transistor current without snubber



Transistor current with snubber

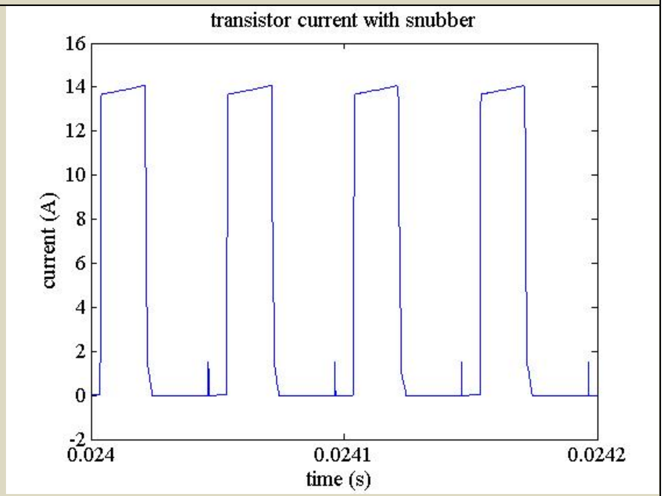
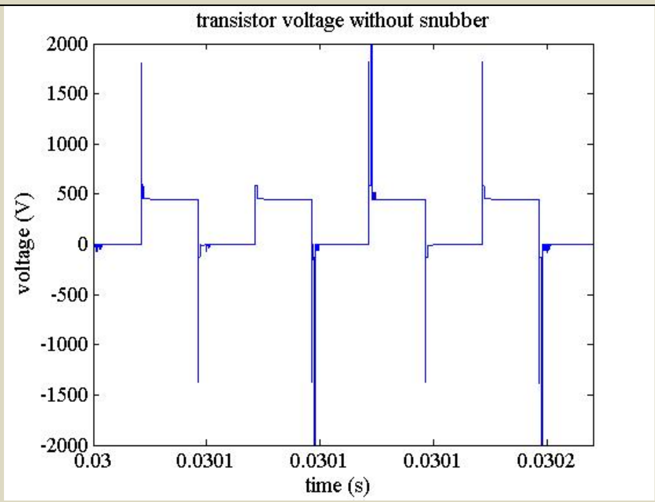


Figure 40 Role of the snubber on the transistor current

Transistor voltage without snubber



Transistor voltage with snubber

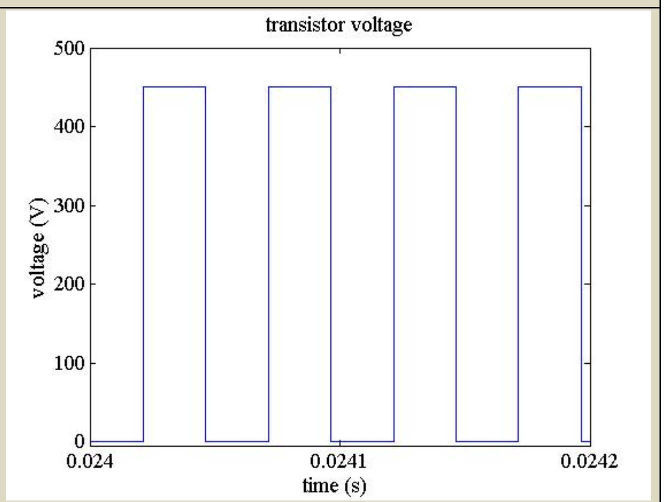


Figure 41 Role of the snubber on the transistor voltage

The snubber reduces stresses during the turn off phases and turn on phases, for the current and the voltage. The snubber reduces switching losses and improves quality of the signals in the on-board charger.

The transformer primary current is shown on the Figure 42. The transformer ratio being equal to one, and the transformer being almost perfect in the simulation, the secondary current has the same waveforms as the primary current. This explains that the secondary voltage of the transformer has the same waveform as the primary voltage (Figure 43).

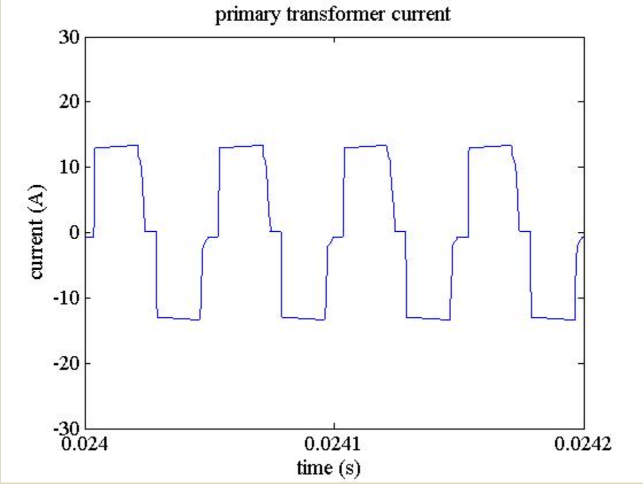
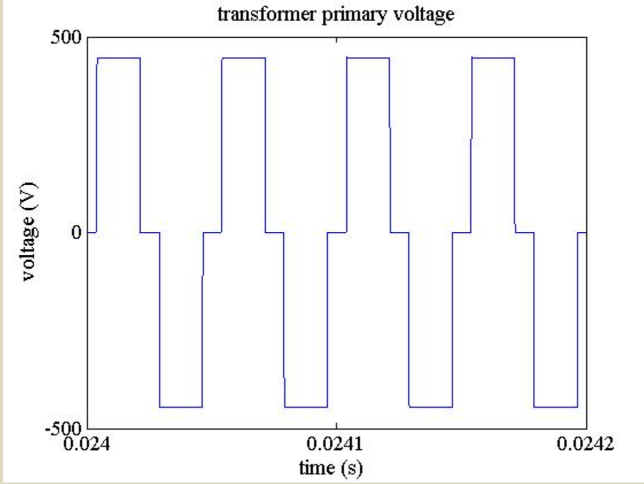
Transformer primary current	Transformer primary voltage
	

Figure 42 Transformer primary current

Figure 43 transformer primary voltage

The current level close to zero should be equal to zero; the difference is due to the transformer inductor. The voltage waveform has the square form which was attending. The variation of the time length of the zero level allows controlling the primary transformer mean voltage.

The Figure 44 shows the current in the inductor filter and the Figure 45 the current in the capacitor filter.

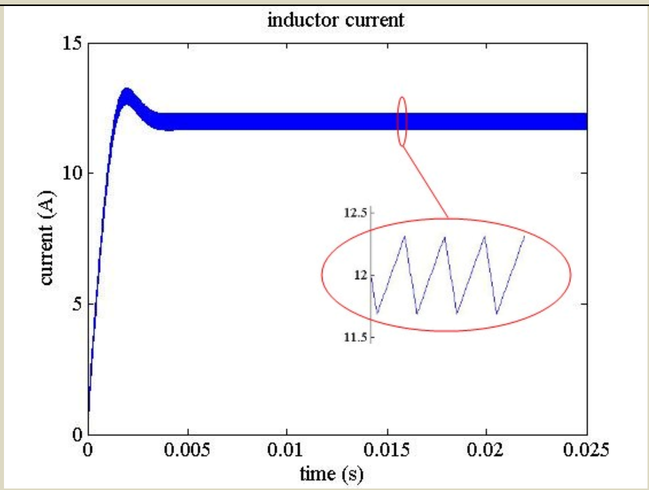
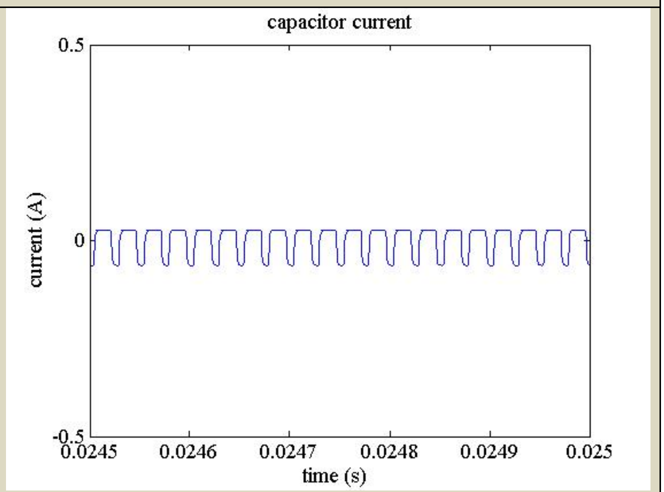
Filter inductor current	Filter capacitor voltage
	

Figure 44 Inductor current

Figure 45 Capacitor current

The current in the filter capacitor is too small. This capacitor current should be the ripple current of the inductor. If the inductor current ripple doesn't go in the capacitor, that means that this current will flow in the battery. This battery current can be viewed on the Figure 46.

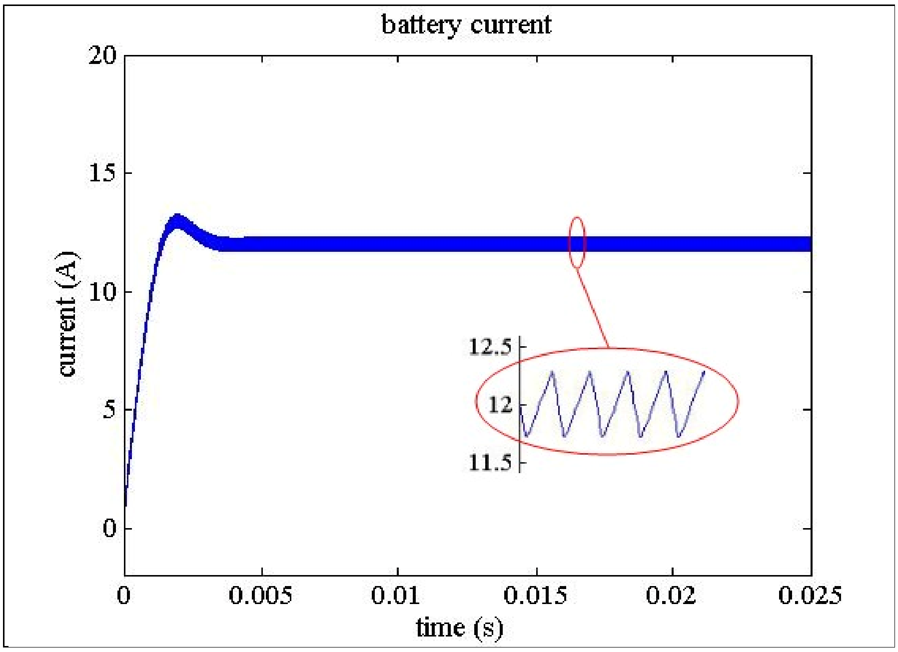


Figure 46 Current in the battery

This figure confirms that the inductor ripple current is flowing in the battery. An EMI filter cab be added to reduce this ripple current but the battery can accept it. The steady state of the battery current is reached after 5 milliseconds.

The Figure 47 shows the battery voltage.

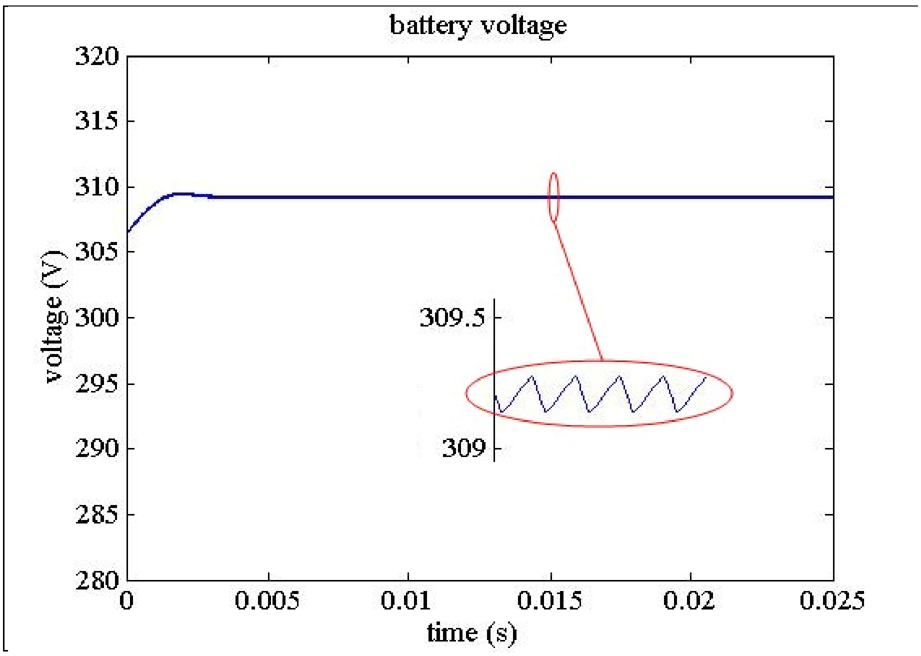


Figure 47 Voltage in the battery

The ripple voltage in the battery in very small (about 0,15V), the steady state is reached before 5 milliseconds.

4) Design

1) Power part

1) Current and voltage value in the on-board charger

In this part, the components are considered ideal, with no losses.

Electrical characteristics 10A:

Input Power P_{in} = 2300W

Input Voltage U_{in} = $U_{in\ rms}\ \sqrt{2}\sin\omega t$

RMS Input Voltage $U_{in\ rms}$ = 230V

Maximum Input Voltage $U_{in\ max}$ = $230\sqrt{2}$ = 325,3V

Frequency f = 50Hz

Power Factor = 1 (cause to the PFC)

PFC Rectifier Bridge:

Input:

$$I_{in} = \frac{P_{in}}{V_{in}\cos\varphi} = \frac{2300}{230*1} = 10A \quad \textbf{(4.1)}$$

$$I_{in\ max} = I_{in}\sqrt{2} = 10\sqrt{2} = 14{,}1\text{A} \quad \textbf{(4.2)}$$

Diode:

$$U_{rms} = \sqrt{\frac{1}{2\pi}\int_{\pi}^{2\pi} \mathrm{V_{in}}^2\sqrt{2}^2\cos^2\theta\, d\theta} \quad \textbf{(4.3)}$$

$$= \mathrm{U_{in}}\sqrt{-1/2} = \text{-162,6V}$$

$$U_{diode\ inv\ max} = \mathrm{U_{in}}\sqrt{2} = \text{- 325,3V} \quad \textbf{(4.4)}$$

$$I_{diode\ rms} = \mathrm{I_{in}}\sqrt{1/2} = 7{,}1\text{A} \quad \textbf{(4.5)}$$

$$I_{diode\ mean} = \frac{1}{2\pi}\int_0^{\pi} \mathrm{I_{in}}\sqrt{2}sin\theta\, d\theta = 4{,}5A \quad \textbf{(4.6)}$$

$$I_{diode\ max} = \mathrm{I_{in}}\sqrt{2} = 14{,}1\text{A} \quad \textbf{(4.7)}$$

Frequency = 50Hz

Output:

$U_{diode\ rms}$ = 230V

$$U_{diode\ mean} = \frac{1}{\pi}\int_0^{\pi} \mathrm{V_{in}}\sqrt{2}sin\theta\, d\theta \quad \textbf{(4.8)}$$

$$= \frac{2\mathrm{V_{in}}\sqrt{2}}{\pi} = 0{,}9\mathrm{V_{in}} = 207\text{V}$$

Frequency = 100Hz

PFC Boost Converter:

Output:

$U_{DC\ link}$ = 450V +/-3%

$$I_{DC\ link} = P_{in}/U_{DC\ link} = 2300/450 = 5{,}11\text{A} \quad \textbf{(4.9)}$$

Electrical characteristics 16A:

Input Power P_{in} = 3700W

Input Voltage U_{in} = $U_{in\ rms}\ \sqrt{2}\sin\omega t$

RMS Input Voltage $U_{in\ rms}$ = 230V

Maximum Input Voltage $U_{in\ max}$ = $230\sqrt{2}$ = 325,3V

Frequency f = 50Hz

Power Factor = 1 (cause to the PFC)

PFC Rectifier Bridge:

Input:

$$I_{in} = \frac{P_{in}}{V_{in}\cos\varphi} = \frac{3700}{230*1} = 16A \quad \textbf{(4.10)}$$

$$I_{in\ max} = I_{in}\sqrt{2} = 16\sqrt{2} = 22{,}6\text{A} \quad \textbf{(4.11)}$$

Diode:

$$U_{rms} = \sqrt{\frac{1}{2\pi}\int_{\pi}^{2\pi} \mathrm{V_{in}}^2\sqrt{2}^2\cos^2\theta\, d\theta} \quad \textbf{(4.12)}$$

$$= \mathrm{V_{in}}\sqrt{-1/2} = \text{-162,6V}$$

$$U_{diode\ inv\ max} = \mathrm{U_{in}}\sqrt{2} = \text{- 325,3V} \quad \textbf{(4.13)}$$

$$I_{diode\ rms} = \mathrm{I_{in}}\sqrt{1/2} = 11{,}3\text{A} \quad \textbf{(4.14)}$$

$$I_{diode\ mean} = \frac{1}{2\pi}\int_0^{\pi} \mathrm{I_{in}}\sqrt{2}sin\theta\, d\theta = 7{,}2A \quad \textbf{(4.15)}$$

$$I_{diode\ max} = \mathrm{I_{in}}\sqrt{2} = 22{,}6\text{A} \quad \textbf{(4.16)}$$

Frequency = 50Hz

Output:

$U_{diode\ rms}$ = 230V

$$U_{diode\ mean} = \frac{1}{\pi}\int_0^{\pi} \mathrm{V_{in}}\sqrt{2}sin\theta\, d\theta \quad \textbf{(4.17)}$$

$$= \frac{2\mathrm{V_{in}}\sqrt{2}}{\pi} = 0{,}9\mathrm{V_{in}} = 207\text{V}$$

Frequency = 100Hz

PFC Boost Converter:

Output:

$U_{DC\ link}$ = 450V +/-3%

$$I_{DC\ link} = P_{in}/U_{DC\ link} = 3700/450 = 8{,}22\text{A} \quad \textbf{(4.18)}$$

Inductor:

U_{ind} = $\mathrm{U_{in}}$ or $\mathrm{U_{in}} - \mathrm{U_{DC\ link}}$ **(4.19)**

= 230V or -170V

I_{rms} = I_{in} = 10A $\Delta I = 20\%$ **(4.20)**

$$L = \frac{V_{DC\ link}}{4\ f_{swi\ max}\ \Delta I} = 1{,}99\mathrm{mH}$$ **(4.21)**

Diode:

U_{diode} = -$\mathrm{U_{DC\ link}}$ = -450V **(4.22)**

$I_{diode\ rms}$ = $I_{inductor\ rms}$ = 10A **(4.23)**

$I_{diode\ max}$ = $\mathrm{I}_{diode\ rms}\sqrt{2}$ = 14,1A **(4.24)**

Transistor:

$U_{transistor}$ = $\mathrm{U_{DC\ link}}$ = 450V **(4.25)**

$I_{transistor\ rms}$ = $I_{inductor\ rms}$ = 10A **(4.26)**

$I_{transistor\ max}$ =$\mathrm{I}_{transistor\ rms}\sqrt{2}$ = 14,1A **(4.27)**

Capacitor:

U_{cap} = $\mathrm{U_{DC\ link}}$ = 450V $\Delta U_{DC\ link} = 3\%$ **(4.28)**

$I_{cap\ max}$ = - $\mathrm{I_{DC\ link}}$ or $\mathrm{I_{ind}}$ - $\mathrm{I_{DC\ link}}$ **(4.29)**

= -5,75A or 8,35A

$$C = \frac{\mathrm{I_{in}}\sqrt{2}\ \mathrm{U_{in}}\sqrt{2}}{2\ \Delta U_{DC\ link}\ \omega\ U_{DC\ link}} = 1{,}36\mathrm{mF}$$ **(4.30)**

Full Bridge Converter:

Output:

According to the battery, the output voltage is somewhere between 290V and 400V.

I_{out} $= \frac{P_{in}}{U_{battery}} = \frac{2300}{290} = 7{,}9A$ **(4.31)**

I_{out} $= \frac{P_{in}}{U_{battery}} = \frac{2300}{400} = 5{,}75A$ **(4.32)**

When the battery is empty, the battery voltage is about 230V (for the 290V's battery).

$I_{out\ max}$ $= \frac{P_{in}}{U_{battery\ min}} = \frac{2300}{230} = 10A$ **(4.33)**

Full Bridge Converter Inverter:

$U_{transistor\ max}$ = 450V

$I_{transistor\ max}$ = 10A

$U_{diode\ max}$ = - 450V

$I_{diode\ max}$ = 10A $f_{switching}$ = 20 kHz

Inductor:

U_{ind} = $\mathrm{U_{in}}$ or $\mathrm{U_{in}} - \mathrm{U_{DC\ link}}$ **(4.34)**

= 230V or -170V

I_{rms} = I_{in} = 16A $\Delta I = 20\%$ **(4.35)**

$$L = \frac{V_{DC\ link}}{4\ f_{swi\ max}\ \Delta I} = 1{,}24\mathrm{mH}$$ **(4.36)**

Diode:

U_{diode} = -$\mathrm{U_{DC\ link}}$ = -450V **(4.37)**

$I_{diode\ rms}$ = $I_{inductor\ rms}$ = 16A **(4.38)**

$I_{diode\ max}$ = $\mathrm{I}_{transistor\ rms}\sqrt{2}$ = 22,6A **(4.39)**

Transistor:

$U_{transistor}$ = $\mathrm{U_{DC\ link}}$ = 450V **(4.40)**

$I_{transistor\ rms}$ = $I_{inductor\ rms}$ = 16A **(4.41)**

$I_{transistor\ max}$ = $\mathrm{I}_{transistor\ rms}\sqrt{2}$ = 22,6A **(4.42)**

Capacitor:

U_{cap} = $\mathrm{U_{DC\ link}}$ = 450V $\Delta U_{DC\ link} = 3\%$ **(4.43)**

$I_{cap\ max}$ = - $\mathrm{I_{DC\ link}}$ or $\mathrm{I_{ind}}$ - $\mathrm{I_{DC\ link}}$ **(4.44)**

= -9,25A or 13,4A

$$C = \frac{\mathrm{I_{in}}\sqrt{2}\ \mathrm{U_{in}}\sqrt{2}}{2\ \Delta U_{DC\ link}\ \omega\ U_{DC\ link}} = 2{,}17\mathrm{mF}$$ **(4.45)**

Full Bridge Converter:

Output:

According to the battery, the output voltage is somewhere between 290V and 400V.

I_{out} $= \frac{P_{in}}{U_{battery}} = \frac{3700}{290} = 12{,}8A$ **(4.46)**

I_{out} $= \frac{P_{in}}{U_{battery}} = \frac{3700}{400} = 9{,}25A$ **(4.47)**

When the battery is empty, the battery voltage is about 230V (for the 290V's battery).

$I_{out\ max}$ $= \frac{P_{in}}{U_{battery\ min}} = \frac{2300}{230} = \frac{3700}{230} = 16A$ **(4.48)**

Full Bridge Converter Inverter:

$U_{transistor\ max}$ = 450V

$I_{transistor\ max}$ = 16A

$U_{diode\ max}$ = - 450V

$I_{diode\ max}$ = 16A $f_{switching}$ = 20 kHz

Full Bridge Converter Transformer:

Transformer ratio = 1

$U_{rms\ max}$ = 450V

$I_{rms\ max}$ = 10A

f = 20 kHz

Full Bridge Converter Rectifier:

RMS Diode Voltage = -450V

Maximum RMS Diode Current = 10A

$f_{switching}$ = 20 kHz

The maximum current in the full bridge converter is 10 Amps. However, in order to keep a constant current during the load and without exceeding 2300W, this current value can be reached.

For a battery having a voltage of 290V, the maximum current will be:

$$I_{out\ max} = \frac{P_{in}}{U_{battery}} = \frac{2300}{290} = 7{,}9A \tag{4.31}$$

And for a battery having a voltage of 400V, the maximum current will be:

$$I_{out\ max} = \frac{P_{in}}{U_{battery}} = \frac{2300}{400} = 5{,}75A \tag{4.32}$$

These values assume a constant current to the load.

For the full bridge semiconductors, the maximum RMS current is 10A. Even if this value won't be reached, this value is the one use to choose the semiconductors in order to protect the components if this value is reached.

Full Bridge Converter Transformer:

Transformer ratio = 1

$U_{rms\ max}$ = 450V

$I_{rms\ max}$ = 16A

f = 20 kHz

Full Bridge Converter Rectifier:

RMS Diode Voltage = -450V

Maximum RMS Diode Current = 16A

$f_{switching}$ = 20 kHz

The maximum current in the full bridge converter is 16 Amps. However, in order to keep a constant current during the load and without exceeding 3700W, this current value can be reached.

For a battery having a voltage of 290V, the maximum current will be:

$$I_{out\ max} = \frac{P_{in}}{U_{battery}} = \frac{3700}{290} = 12{,}75A \tag{4.46}$$

And for a battery having a voltage of 400V, the maximum current will be:

$$I_{out\ max} = \frac{P_{in}}{U_{battery}} = \frac{3700}{400} = 9{,}25A \tag{4.47}$$

These values assume a constant current to the load.

For the full bridge semiconductors, the maximum RMS current is 16A. Even if this value won't be reached, this value is the one using to choose the semiconductors in order to protect the components if this value is reached.

2) Losses in the semiconductors

Losses in the diode

The losses in a diode consist of the conduction losses and switching losses.

The conduction losses depend on the forward voltage drop of the diode, the forward current and the internal resistance of the diode.

$$P_{cond} = V_F I_F \tag{4.49}$$

Where V_F is the voltage drop of the diode, I_F the current in the diode.

In the freewheeling diode, the conduction losses become:

$$P_{cond} = V_{CE} I_F \tag{4.50}$$

Where V_{CE} is the voltage drop of the IGBT transistor, I_F the courant in the diode.

The switching losses consist of the turn off and turn on, and reverse recovery losses. At low frequency these losses can be ignored with the use of fast diode due to their very short reverse recovery time. These losses can be calculated using the following expression:

$$P_{swi} = f_{swi} E_{off} \tag{4.51}$$

Where f_{swi} is the switching frequency, E_{off} is the energy dissipated during the recovery time. E_{off} can be find in the component datasheet.

The switching losses are in principally the losses during the reverse recovery time so these losses can be determined as:

$$P_{swi} \simeq P_{rr} = f_{swi} \left(\frac{V_r\, I_r}{2} \right) t_b = f_{swi} E_{rr} \tag{4.52}$$

Where f_{swi} is the switching frequency, V_r is the peak reverse recovery voltage, I_r is the peak reverse recovery current, t_b the snap off time, and E_{rr} the energy during the reverse recovery time.

The losses in the diode are:

$$P_{diode} = P_{cond} + P_{swi} \tag{4.53}$$

Losses in the transistor

The losses in a transistor consist of the conduction losses and switching losses.

IGBT

The IGBT conduction losses depend on the saturation voltage drop of the transistor, the forward current and the internal resistance of the transistor.

$$P_{cond} = V_{CE} I_C \tag{4.54}$$

Where V_{CE} is the saturation voltage drop of the transistor, I_c the current in the transistor.

The IGBT switching losses can be calculated using the following relation:

$$P_{swi} = f_{swi}(E_{on} + E_{off}) \quad \textbf{(4.55)}$$

Where f_{swi} is the switching frequency, E_{off} is the energy loss during the off state switching, E_{on} is the energy loss during the on state switching. $E_{off}$and E_{on} can be found in the component datasheet.

MOSFET

The MOSFET conduction losses depend on the saturation voltage drop of the transistor, the forward current and the internal resistance of the transistor.

$$P_{cond} = r_{DS(on)}{I_D}^2 \quad \textbf{(4.56)}$$

Where I_D the courant in the transistor, and r_{DS} the internal resistance of the transistor during the on state.

The MOSFET switching losses can be calculated using the following expression:

$$P_{swi} = \left(f_{swi}\, C_{rss}\, V_{DS}^2\, I_D\right)/I_G \quad \textbf{(4.57)}$$

Where $f_{swi}$is the switching frequency, C_{rss} is the reverse-transfer capacitance, $V_{DS}$is the transistor voltage,I_D the courant in the transistor, and I_G is the source current at the turn-on threshold.

Or:

$$P_{swi} = f_{swi}\ V_{DS}\, I_D\, t_{swi} \quad \textbf{(4.58)}$$

Where $f_{swi}$is the switching frequency, $V_{DS}$is the transistor voltage, I_D the current in the transistor, and t_{swi} is the switching total time.

The losses in the transistor are:

$$P_{transistor} = P_{cond} + P_{swi} \quad \textbf{(4.59)}$$

3) Losses in the transformer

The losses in the transformer consist of the losses in the windings and also the losses in the core material.

The losses in the windings with a phase shift control can be calculated using the winding resistances:

$$P_w = R_{prim}\, i_{prim}^{2} + R_{sec}\, i_{sec}^{2} \quad \textbf{(4.60)}$$

Where R_{prim} is the primary winding resistance, i_{prim} is the primary winding current, R_{sec} is the secondary winding resistance, i_{sec} is the secondary winding current.

With a duty cycle control, these losses become:

$$P_w = D(R_{prim}\, i_{prim}^{2} + R_{sec}\, i_{sec}^{2}) \quad \textbf{(4.61)}$$

Where D is the duty cycle, R_{prim} is the primary winding resistance, i_{prim} is the primary winding current, R_{sec} is the secondary winding resistance, i_{sec} is the secondary winding current.

The losses in the core material need the value of the peak magnetic field in the core to be calculated:

$$B_{max} = V_{prim} \frac{D}{2f_{swi}} \frac{1}{2N_{prim}A_{core}} \quad \textbf{(4.62)}$$

Where V_{prim} is the primary voltage, D the duty cycle, f_{swi} is the switching frequency, N_{prim} is the number of primary turns and A_{core} the area of the core.

The losses in the core material are the following:

$$P_{core} = K_1 \, f_{swi}^{K2} \, B_{max}^{K3} \, V_{core} \quad \textbf{(4.63)}$$

Where V_{core} is the core voltage and K_1, K_2, K_3 are coefficients depending on the core material.

The losses in the transformer are:

$$P_{transf} = P_w + P_{core} \quad \textbf{(4.64)}$$

4) Losses in the transformer and semiconductor on-board charger

The losses are the sum of the losses in the diode input bridge rectifier, diode and IGBT boost converter, IGBT and freewheeling diode inverter, transformer and diode output bridge rectifier. The losses in the on-board charger differ according to the full bridge converter control method.

The losses in the semiconductors and transformer with the duty cycle control method are the following:

$$P_{OBC} = 2(V_F I_{Fav}) + f_{swi}E_{off} + V_F I_{Fav} + f_{swi}\left(E_{on} + E_{off}\right) + V_{CE}I_{Cav} + 2D(V_{CE}I_{Cav}) + D\left(R_{prim}\, i_{prim}^{2} + R_{sec}\, i_{sec}^{2}\right) + K_1\, f_{swi}^{K2}\, B_{max}^{K3}\, V_{core} + 2D(V_F I_{Fav}) + 4(1-D)\left(\frac{V_F I_{Fav}}{2}\right) + 2\left(f_{swi}E_{off}\right) + 4(f_{swi}E_{off\ i=\frac{I}{2}}) \quad \textbf{(4.65)}$$

The losses in the freewheeling diodes are assumed to be small with the duty cycle control.

The losses in the semiconductors and transformer with the phase shift control method are the following:

$$P_{OBC} = 2(V_F I_{Fav}) + f_{swi}E_{off} + V_F I_{Fav} + f_{swi}\left(E_{on} + E_{off}\right) + V_{CE}I_{Cav} + 2D(V_{CE}I_{Cav}) + (1-D)(V_{CE}I_{Cav}) + (1-D)(V_{CE}I_{Fav}) + \left(R_{prim}\, i_{prim}^{2} + R_{sec}\, i_{sec}^{2}\right) + K_1\, f_{swi}^{K2}\, B_{max}^{K3}\, V_{core} + 2(V_F I_{Fav}) + 2\left(f_{swi}E_{off}\right) \quad \textbf{(4.66)}$$

5) Snubbers

Previously, two kinds of snubbers have been studied (see: Snubber). In this full bridge converter, snubbers for the transistors are capacitor circuits (Figure 48). Theses circuits allow limiting stresses during the turn off phases and also limited dv/dt across the transistor.

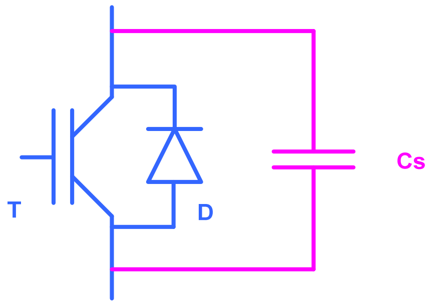


Figure 48 Snubber capacitor

The role of the capacitor is to absorb current that otherwise would go through the transistor and in this way reducing loss stress for the transistor during the turn off phase. The capacitor value can be obtained by the following relation:

$$C_s = \frac{I_O t_{fi}}{2V_d} \tag{4.67}$$

Where I_O is the transistor current and $\mathrm{I_O}$ = 12,75A, t_{fi} is the current fall time, V_d is the transistor supply voltage and $\mathrm{V_d}$ = 400V.

To choose the current fall time, a *di/dt* value has to be decided. According to the component datasheet, 100A/µs is a good value.

$$t_{fi} = \frac{12{,}75}{100} * 1.10^{-6} = 0{,}13.10^{-6}s \tag{4.68}$$

The snubber capacitor value is:

$$\mathrm{C_s} = \frac{12{,}75 \ * 0{,}13.10^{-6}}{2 * 400} = 2{,}07n\mathrm{F}$$

6) Transformer

Transformers are not commercially available in a wide range of properties and they are usually designed for a particular application. In this chapter, a transformer is designed.

Design of the transformer

<u>Parameters:</u>

Input current $I_{pri\,max} = 16A$

Input voltage $V_{pri\,max} = 450V$

Frequency $f = 20kHz$

Turns ratio $n = 1$

<u>Volt-amp rating of the transformer:</u>

$$S = U * I = 3700\,VA \quad \textbf{(4.69)}$$

<u>Core material, shape and size:</u>

The core used for this transformer is two ETD core put together. The primary and secondary turns are wounded on the centre leg. Figure 49 shows an ETD core transformer.

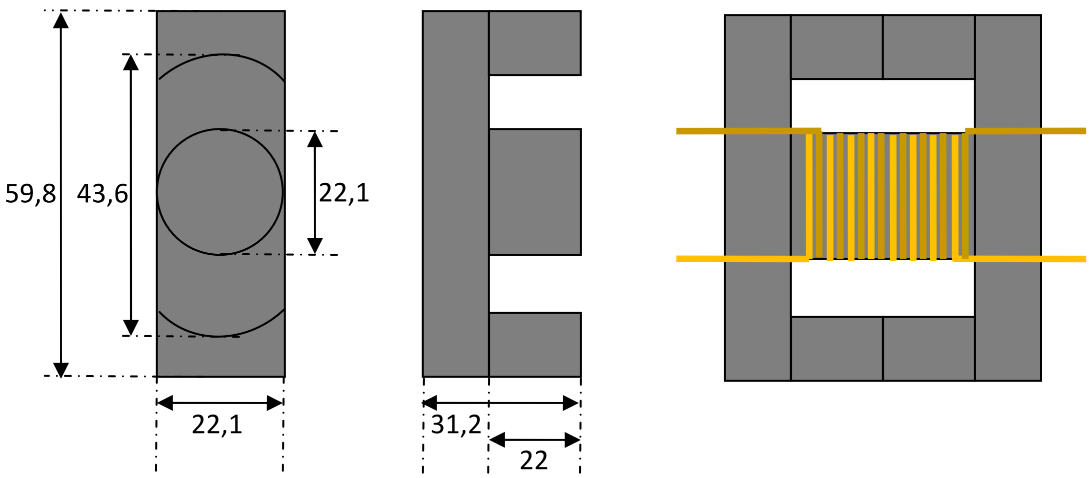


Figure 49 Size and shape of the core material

This core is built with ferrite material. ETD cores in many different ferrite materials and core sizes are provided by several manufacturers. The core chosen is called ETD 59/31/22, from the supplier Elfa, where the different numbers are dimensions of the core. Table 1 shows the specifications of the core are presented.

Manufacturer	Elfa
Reference	ETD 59/31/22
Volume of the ferrite material (mm^3)	51200
Volume of the core (mm^3)	82500
Cross sectional area(mm^2)	368
Mass (g)	520
Inductance index (nH/N²)	5500
$\widehat{B}$sat(T)	0,410

Table 1 Transformer core parameters

$$A_{core} = \pi * (11{,}05.10^{-3})^2 = 3{,}68\, cm^2 \quad \textbf{(4.70)}$$

The maximum temperature of the core will be $T_s = 100°C$ and the ambient temperature is assumed to be about $T_a = 40°C$. The maximum temperature rise of the transformer core is:

$$T_{rise} = T_s - T_a = (100 + 273{,}16) - (40 + 273{,}16) = 60°K \quad \textbf{(4.71)}$$

The thermal resistance of the core can be found in the datasheet; this value is $R_{th} = 1{,}5K/W$. The maximum allowed total power loss in the transformer is:

$$P_{core\ max\ loss} = \frac{T_{rise}}{R_{th}} = \frac{60}{1{,}5} = 40W \quad \textbf{(4.72)}$$

The number of turns is needed for the primary side can be calculated by the following expression:

$$N_{pri} = \frac{\hat{V}_{pri}}{A_{core}\omega\hat{B}_{core}} = \frac{400}{3{,}68.10^{-4} * 2\pi * 20.10^3 * 0{,}410} = 21{,}09\ turns \quad \textbf{(4.73)}$$

The turn ratio of the transformer is 1 so the number of secondary turns is equal to the number of primary turns.

$$N_{sec} = N_{pri} = 21 \quad \textbf{(4.74)}$$

The windings of the core create an inductance which is called the magnetization inductance. The magnetization inductance has to be charged by a current in order for the transformer to work.

$$L_m = 2A_L * {N_{pri}}^2 = 2 * 5500.10^{-9} * 21^2 = 4{,}86mH \quad \textbf{(4.75)}$$

The wire chosen, for the windings of this transformer, is a Litz wire. This wire allows reducing the eddy current and the skin effect. This wire has a cross sectional area of 0,94mm² (largest size provided by ELFA). This wire can accept a current of 3,36A.

The maximum current in the transformer is 12,75A so, as a result, several wires have to be paralleled:

$$\frac{I_{rms\,max}}{I_{wire\,max}} = \frac{12{,}75}{3{,}36} = 3{,}8 \rightarrow 4\,wires \tag{4.76}$$

The total wire area is:

$$A_{wire} = 4 * 0{,}94 = 3{,}76mm^2 \tag{4.77}$$

The wire diameter is:

$$d_{wire} = 2\sqrt{\frac{A_{wire}}{\pi}} = 2\sqrt{\frac{3{,}76}{\pi}} = 2{,}18mm \tag{4.78}$$

The total wire area is:

$$A_{wire\,total} = 3{,}76.10^{-3} * (21 + 21) = 1{,}58\,cm^2 \tag{4.79}$$

The maximum area wire possible is:

$$A_{wire\,possible} = (8{,}1 * 44).10^{-3} = 3{,}56\,cm^2 \tag{4.80}$$

That gives that there is no problem for the primary and secondary windings to fit on the core.

Verification:

The maximum power in a transformer can be calculated using the following expression:

$$P_{max} = \frac{N_{pri} * A_{core} * \omega * \hat{B}}{\sqrt{2}} * J_{rms} * A_{cu\,pri} \tag{4.81}$$

$$P_{max} = \frac{21 * 3{,}68.10^{-4} * 2\pi * 20000 * 0{,}410}{\sqrt{2}} * 3{,}57 * 3{,}76 = 3779W$$

The maximum power admissible by the transformer is higher than the input power so the design and the choice of the core material of the transformer is correct.

Losses in the transformer

Wire losses:

Length of the wire:

$$l = N * (\pi * d_{core}) + N * (\pi * (d_{core} + d_{wire}) \tag{4.82}$$

$$l = 21 * (\pi * 22{,}1.10^{-3}) + 21 * (\pi * (22{,}1.10^{-3} + 2{,}18.10^{-3})$$

$$l = 3{,}60m$$

Resistance of the wire:

$$R_W = \frac{\rho_{cu}.l_W}{A_W} = \frac{1{,}7.10^{-8} * 3{,}6}{3{,}76.10^{-6}} = 0{,}0163\Omega \tag{4.83}$$

Losses in the wire:

$$P_W = R_W * {I_L}^2 = 0{,}0163 * 12{,}75^2 = 2{,}65W \tag{4.84}$$

Core losses:

The core losses can be determined thanks to the datasheet of the core. The core losses are about $175kW/m^3$

The volume of the core can be finding in the datasheet and its value is $51200mm^3$

$$P_{core} = losses\ density * volume = 155.10^3 * 51{,}2.10^{-6} = 8{,}96W \quad \textbf{(4.85)}$$

Transformer losses:

$$P_{total} = P_W + P_{core} = (2 * 2{,}65) + 9 = 11{,}65W \quad \textbf{(4.86)}$$

7) Choice of the semiconductor

Choice of the diode

Rectifier Soft Recovery:

Type	Manufacturer Reference	Fairchild Semiconductor RHRP3060	Fairchild Semiconductor RURP3060
Rectifier Soft Recovery	$I_{f\ average}$ (A)	30	30
	$V_{f\ @\ I=12,7}$ (V)	1,15	0,95
	$V_{f\ @\ I=16}$ (V)	1,25	1
	V_{rrm} (V)	600	600
	Q_{rr} (nC)	350	?
	t_{rr} (ns)	35	45
	t_b (ns)	16	18
	Price (kr)	87,67	52,50
	Conduction losses $@\ I = 12,7$	14,6	12,06
	Switching losses	1,4	?
	Total losses $@\ I = 12,7$	16	
	Conduction losses $@\ I = 16$	20	16
	Switching losses	1,4	?
	Total losses $@\ I = 16$	21,4	

Rectifier Fast Recovery:

Type	Manufacturer/Reference	Vishay Rectifier 30ETH06PBF	IXYS Semiconductor DSEP29-06A
Rectifier Fast Recovery	$I_{f\ average}$ (A)	30	30
	$V_{f\ @\ I=12,7}$ (V)	1,2	1,15
	$V_{f\ @\ I=16}$ (V)	1,25	1,2
	V_{rrm} (V)	600	600
	Q_{rr} (nC)	250	300
	t_{rr} (ns)	31	35
	t_b (ns)	13*	14*
	Price (kr)	46,00	25,52
	Conduction losses $@\ I = 12,7$	15,24	14,6
	Switching losses	1	1,2
	Total losses $@\ I = 12,7$	16,24	15,8
	Conduction losses $@\ I = 16$	20	19,2
	Switching losses	1	1,2
	Total losses $@\ I = 16$	21	20,4

*Approximate value (not come from datasheet)

For the input rectifier bridge diode and the full bridge freewheeling diode, they are no switching losses. For the rectifier bridge, it is due to the low frequency and for the freewheeling diodes, it is due to the snubbers. The diode having the low conduction losses is *Fairchild Semiconductor RURP3060* so this diode is used in these converters. For the PFC diode and the output rectifier bridge, where both conduction and switching losses are considered, the diode used is *IXYS Semiconductor DSEP29-06A.*

Choice of the transistor

IGBT Transistor

Type	Manufacturer/Reference	International rectifier IRG4BC40UPBF	Infineon IKW30N60T
IGBT	$I_{c\ continuous\ a\ max}$ (A)	20	30
	$V_{ce\ sat\ @\ I=12,7}$ (V)	1,6	1,25
	$V_{ce\ sat\ @\ I=16}$ (V)	1,65	1,4
	V_{ces} (V)	600	600
	E_{on} $@\ I = 12{,}7$	0,85 mJ	0,4 mJ
	E_{off} $@\ I = 12{,}7$		0,65 mJ
	E_{on} $@\ I = 16$	1,1 mJ	0,55 mJ
	E_{off} $@\ I = 16$		0,75 mJ
	Price (kr)	78,03	53,29
	Conduction losses $@\ I = 12{,}7$	20,32 W	15,88 W
	Switching losses$@\ I = 12{,}7$	17 W	21 W
	Total losses $@\ I = 12{,}7$	37,3 W	36,88 W
	Conduction losses $@\ I = 16$	26,4 W	22,4 W
	Switching losses$@\ I = 16$	22 W	26 W
	Total losses $@\ I = 16$	48,4 W	48,4 W

MOSFET Transistor

Type	Manufacturer/Reference	Vishay Rectifier IRFPS30N60KPBF	STMicroelectronics STW26NM60
MOSFET	$I_{d\ continuous}$ (A)	30	30
	V_{ds} (V)	600	600
	r_{ds} (Ω) $@\ I = 12{,}7$	0,13*	0,12
	r_{ds} (Ω) $@\ I = 16$	0,16*	0,145
	t_{swi} (ns) $@\ I = 12{,}7$	108	89
	t_{swi} (ns) $@\ I = 16$	136	112
	Price (kr)	131,61	100,93
	Conduction losses $@\ I = 12{,}7$	20,97	19,35
	Switching losses$@\ I = 12{,}7$	10,97	9,04
	Total losses $@\ I = 12{,}7$	31,94	28,39
	Conduction losses $@\ I = 16$	40,96	37,12
	Switching losses $@\ I = 16$	17,41	14,34
	Total losses $@\ I = 16$	58,37	51,46

*Approximate value

For the PFC transistor, the conduction losses and switching losses are considered. The transistor having less loss is Infineon IKW30N60T so this transistor is used in the PFC part. For the full bridge transistors, only the conduction losses are considered, the switching losses are assumed to be small thanks to the snubbers. In this converter, it’s also the Infineon IKW30N60T which is used.

8) Losses in the passive components

Inductor

Wire losses

$$P_W = R_W * {I_L}^2 \tag{4.87}$$

With R_W is the wire resistance and I_L the current in the inductance.

$$R_W = \frac{\rho_{cu} \cdot L_W}{A_W} \tag{4.88}$$

With ρ_{cu} is the resistivity of the wire material (copper), L_W the length of the wire and A_W the copper area of the wire.

For a copper wire, the resistivity is:

$\rho_{cu} = 1{,}7.10^{-8}\Omega.m$

Core losses

The core losses can be found thanks to the datasheet. These losses vary according to AC flux density (in Gauss), the frequency and the volume of the core.

The AC flux density depends on the current ripple and can be calculated thanks to the following expression:

$$\hat{B} = \frac{L * \Delta\hat{\imath}_L}{N * A} \tag{4.89}$$

Where L is the inductance value, $\Delta\hat{\imath}_L$ is the current ripple, N is the number of turns and A is the inductor core area.

The magnetic field is in Tesla (T) and 1 Tesla = 10^4 Gauss

Capacitor

$$P_{losses} = {I_c}^2 * (ESR) \tag{4.90}$$

Where I_c is the current in the capacitor. ESR can be found in the capacitor datasheet.

9) Choice of the PFC components

PFC inductor

Parameters:

Inductance value: $L = 1{,}24\ mH$

Rated peak current: $\hat{I} = 17{,}6A$

Rated RMS current: $I_{rms} = 16A$

Operating frequency: $f = 20\ kHz$

Stored energy value:

$$L\hat{I}I_{rms} = 1{,}24.10^{-3} * 17{,}6 * 16 = 0{,}349\ HA^2 \quad \textbf{(4.91)}$$

Core material and shape:

The shape of the core is decided to be a toroid shaped core. This is a very common shape provided by many core manufacturers. In Figure 50 an inductor with a toroids shaped core is shown. The core material used is an iron powder material. Iron powder cores have larger saturation flux density than ferrite cores but larger core losses. The frequency is not high enough to give important losses. Table 2 gives the inductor core parameters.

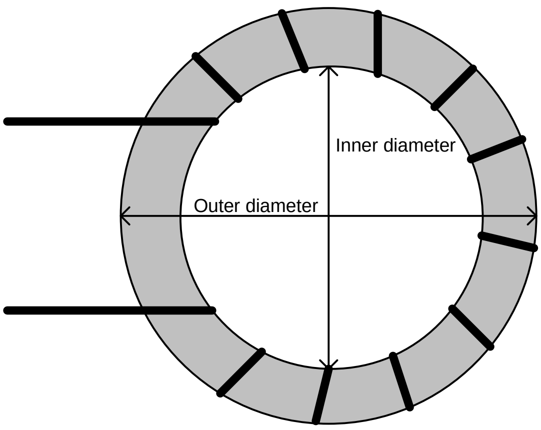


Figure 50 Shape of the inductor

Manufacturer	Amidon
Reference	T300A-26
Inner diameter (mm)	48,9
Outer diameter (mm)	77,4
Height (mm)	25,4
Cross area (mm²)	362
Volume (mm^3)	70991
Weight (kg)	0,497
B_{sat}(T)	1,5
Inductance index (µH/100 turns)	1600

Table 2 Core inductor value

Wire:

Maximum current: $\hat{\imath}_{L\ max} = 17{,}6A$

The current density in the wire is 6A/mm²

Cross sectional area of the copper wire:

$$A_{wire} = \frac{\hat{\imath}_{L\,max}}{current\ density} = \frac{17{,}6}{6} = 2{,}93\ mm^2 \tag{4.92}$$

From the cross sectional area of the copper wire the diameter of the wire can be found:

$$d = 2\sqrt{\frac{A_{wire}}{\pi}} = 2\sqrt{\frac{3}{\pi}} = 1{,}96mm \tag{4.93}$$

The number of turn needed for the core:

$$N = \frac{L * 100}{A_L} = \frac{1{,}24.\,10^3 * 100}{1600} = 77{,}5\ turns \tag{4.94}$$

Number of turn possible for this core:

$$N_{possible} = \frac{\pi * d_{inner\ core}}{d_{wire}} = \frac{\pi * 48{,}9}{1{,}96} = 78{,}4\ turns \tag{4.95}$$

Magnetic field in this core:

$$\hat{B} = \frac{L * \hat{\imath}_L}{N * A} = \frac{1{,}24.\,10^{-3} * 17{,}6}{78 * 3{,}62.\,10^{-4}} = 0{,}778T \tag{4.96}$$

Verification:

$$\hat{\Phi} = \hat{B}_{core} * A_{core} = 0{,}778 * 3{,}62.\,10^{-4} = 2{,}82.\,10^{-4}Wb \tag{4.97}$$

$$L = \frac{N\hat{\Phi}}{\hat{I}} = \frac{78 * 2{,}82.\,10^{-4}}{17{,}4} = 1{,}24mH \tag{4.98}$$

Losses in the boost inductor

Wire losses:

Length of the wire:

$$l = N * 2 * \big(height + (outer\ diameter - inner\ diameter)\big) \tag{4.99}$$

$$= 79 * 2\big(25{,}4 + (77{,}4 - 48{,}9)\big)$$

$$l = 8{,}52m$$

Resistance of the wire:

$$R_W = \frac{\rho_{cu}.\,l_W}{A_W} = \frac{1{,}7.\,10^{-8} * 8{,}52}{3.\,10^{-6}} = 0{,}048\Omega \tag{4.100}$$

Losses in the wire:

$$P_W = R_W * {I_L}^2 = 0{,}048 * 16^2 = 12{,}3W \tag{4.101}$$

Core losses:

$$\widehat{B_{AC}} = \frac{L * \Delta\hat{\imath}_L}{N * A} = \frac{1{,}24.10^{-3} * 1{,}6}{78 * 3{,}62.10^{-4}} = 0{,}070 \text{ T or } 702{,}6 \text{ G} \quad \textbf{(4.102)}$$

The datasheet gives, for this flux density at a 20 kHz frequency, a loss value of $140mW/cm^3$

Core volume:

$$V_{core} = \left(\pi * (\frac{diameter_{outer}}{2})^2 - \pi * (\frac{diameter_{inner}}{2})^2\right) * height \quad \textbf{(4.103)}$$

$$V_{core} = \left(\pi * (\frac{77{,}4}{2})^2 - \pi * (\frac{48{,}9}{2})^2\right) * 25{,}4 = 71{,}8cm^3$$

Losses in the core:

$$P_{core} = 0{,}140 * 71{,}8 = 10{,}1W \quad \textbf{(4.104)}$$

Losses in the inductor:

$$P_{inductor} = P_{core} + P_W = 10{,}1 + 12{,}3 = 22{,}4W \quad \textbf{(4.105)}$$

Capacitor

The capacitor will be an aluminium electrolytic capacitor for the PFC DC link

Type	Manufacturer/Reference	EPCOS B43584
Aluminium Electrolytic	Capacitor value (µF)	1800
	Voltage rating (V)	500
	ESR (mΩ)	59
	Weight (kg)	0,44
	Volume (dm^3)	0,34
	I_{rms} (A)	7,9

Losses in PFC DC link capacitor

$$P_{losses} = {I_c}^2 * (ESR) = 1{,}6^2 * (0{,}059) = 0{,}15W \quad \textbf{(4.106)}$$

10) Output filter

The Figure 51 shows a low pass filter.

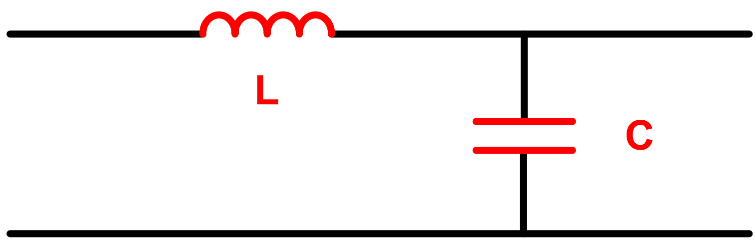


Figure 51 output filter

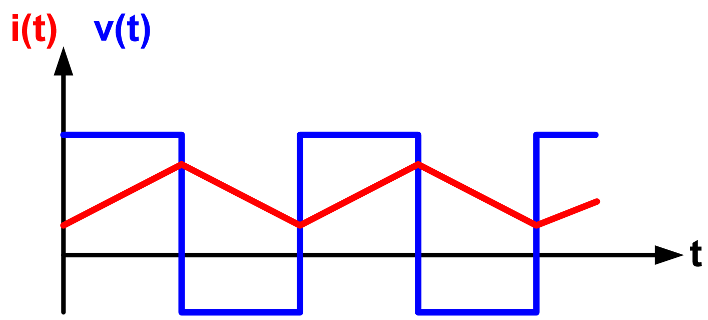
Inductor value

The following relation is used to calculate the value of the inductor:

$$V_L(t) = L\frac{di_L}{dt} \tag{4.107}$$

Where V_L is the inductor voltage and i_L the current in the inductor.

Figure 52 shows the current and voltage waveforms in the inductor.



frequency: f=20 kHz

period: t=$5.10^{-5}s$

half period: t=$2,5.10^{-5}s$

Figure 52 Current and voltage in the inductor

The current ripple is fixed at about 15% of the RMS inductor current. For a battery voltage of 400V, the inductor current is 9,25A.

$$\Delta I = 15\%\ I_{inductor} = 0,15 * 9,25 = 1,39A \tag{4.108}$$

For a battery of 290V, the inductor current is 12,75A. The current ripple of 15% represents 1,91Amps.

For a battery of 290V:

$$L_{min} = \frac{V_{L\,max}}{\frac{di_L}{dt}} = \frac{450 - 0,75 * 290}{\frac{1,91}{2,5.10^{-5}}} = 3,05mH \tag{4.109}$$

For a battery of 400V:

$$L_{max} = \frac{V_{L\,max}}{\frac{di_L}{dt}} = \frac{450 - 0,75 * 400}{\frac{1,39}{2,5.10^{-5}}} = 2,7mH \tag{4.110}$$

The inductor has to be designed for a value of 3,05mH. With this value the maximum ripple current will be 15%.

Capacitor value

The following relation is used to calculate the value of the capacitor:

$$i_C(t) = C\frac{dV_C}{dt} \tag{4.111}$$

Where V_c is the capacitor voltage and i_c the current in the capacitor.

The current in the capacitor is the ripping current of the inductor so the value of this current is between 1,3A and 1,9A. The capacitor voltage is identical to the battery voltage so its value is varying according to the battery voltage. The capacitor voltage ripple is fixed at 5% and the frequency is 20 kHz.

For a battery of 290V:

$$C_{max} = \frac{I_c}{\frac{dV_c}{dt}} = \frac{1{,}91}{\frac{0{,}05 * 290}{2{,}5.10^{-5}}} = 3{,}3\mu F \tag{4.112}$$

For a battery of 400V:

$$C_{min} = \frac{I_c}{\frac{dV_c}{dt}} = \frac{1{,}39}{\frac{0{,}05 * 400}{2{,}5.10^{-5}}} = 1{,}74\mu F \tag{4.113}$$

An appropriate value of the capacitance is thus $3{,}3\mu F$.

Cutoff frequency

The cutoff frequency is calculated using the following relation:

$$\omega_c = \sqrt{\frac{1}{LC}} \tag{4.114}$$

Where ω_c is the cutoff pulsation, L is the inductor value in Henries and C the capacitor value in Farads.

The cutoff frequency is:

$$\omega_c = \sqrt{\frac{1}{LC}} = \sqrt{\frac{1}{3{,}05.10^{-3} * 3{,}3.10^{-6}}} = 9968\, rad.s^{-1} \tag{4.115}$$

$f_{c\,min} = 1586Hz$

The cutoff frequency is about 1600Hz, it can be calculated more precisely when the capacitor and inductor will be choose. This frequency represents 8% of the switching frequency, this value is correct.

11) Choice of the filter components

Inductor

Parameters:

Inductance value: $L = 3\ mH$

Rated peak current: $\hat{I} = 14\ A$

Rated RMS current: $I_{rms} = 12{,}75\ A$

Operating frequency: $f = 20\ kHz$

Stored energy value:

$$L\hat{I}I_{rms} = 3.10^{-3} * 14 * 12{,}75 = 0{,}536\ HA^2 \quad \textbf{(4.116)}$$

Choice of the core material and shape:

The shape of the core is decided to be a toroids shaped core. This is a very common shape provided by many core manufacturers. The core material used is an iron powder material. Iron powder cores have larger saturation flux density than ferrite cores but larger core losses. The frequency is not enough high to give important losses. Table 3 gives the core inductor parameters.

Manufacturer	Amidon
Reference	T300A-26
Inner diameter (mm)	48,9
Outer diameter (mm)	77,4
Height (mm)	25,4
Cross area (mm^2)	358
Volume (mm^3)	70991
Weight (kg)	0,497
$\mathrm{B_{sat}}(\mathrm{T})$	1,5
Inductance index (nH/N^2)	131

Table 3 Inductor core value

Wire:

Maximum current: $\hat{\imath}_{L\ max} = 14A$

The current density in the wire is $6A/mm^2$

Cross sectional area of the copper wire:

$$A_{wire} = \frac{\hat{\imath}_{L\ max}}{current\ density} = \frac{14}{6} = 2{,}33\ mm^2 \quad \textbf{(4.117)}$$

From the cross sectional area of the copper wire the diameter of the wire can be found:

$$d = 2\sqrt{\frac{A_{wire}}{\pi}} = 2\sqrt{\frac{2{,}5}{\pi}} = 1{,}78mm \quad \textbf{(4.118)}$$

Number of turns need for the core:

$$N = \sqrt{\frac{L}{A_L}} = \sqrt{\frac{3.10^6}{131}} = 136{,}33\ turns \tag{4.119}$$

Number of turns possible with this core:

$$N_{possible} = \frac{\pi * d_{inner\ core}}{d_{wire}} = \frac{\pi * 48{,}9}{1{,}78} = 86\ turns \tag{4.120}$$

The turns needed are higher than the maximum possible turns on one layer; the solution is to have a second layer of wire on the core.

Magnetic field in this core:

$$\hat{B} = \frac{L * \hat{\imath}_L}{N * A} = \frac{3.10^{-3} * 14}{137 * 3{,}58.10^{-4}} = 0{,}857T \tag{4.121}$$

The magnetic field in this core is lower than the saturation magnetic field.

Losses in the output filter inductor

<u>Wire losses:</u>

Length of the wire:

$$l = N * 2 * \big(height + (outer\ diameter - inner\ diameter)\big) \tag{4.122}$$

$$= 137 * 2 * (25{,}4 + (77{,}4 - 48{,}9))$$

$l = 14{,}77m$

Resistance of the wire:

$$R_W = \frac{\rho_{cu}.L_W}{A_W} = \frac{1{,}7.10^{-8} * 14{,}77}{2{,}5.10^{-6}} = 0{,}100\Omega \tag{4.123}$$

Losses in the wire:

$$P_W = R_W * {I_L}^2 = 0{,}1 * 12{,}75^2 = 16{,}3W \tag{4.124}$$

<u>Core losses:</u>

$$\widehat{B_{AC}} = \frac{L * \Delta\hat{\imath}_L}{N * A} = \frac{3.10^{-3} * ((1{,}91 + 1{,}39)/2)}{137 * 3{,}58.10^{-4}} = 0{,}1 \text{ T or } 100{,}1 \text{ G} \tag{4.125}$$

The datasheet gives, for this flux density at a 20 kHz frequency, a loss value of $300mW/cm^3$

Core volume:

$$V_{core} = \left(\pi * (\frac{diameter_{outer}}{2})^2 - \pi * (\frac{diameter_{inner}}{2})^2\right) * height \tag{4.126}$$

$$V_{core} = \left(\pi * (\frac{77{,}4}{2})^2 - \pi * (\frac{48{,}9}{2})^2\right) * 25{,}4 = 71{,}8cm^3$$

Losses in the core:

$$P_{core} = 0{,}300 * 71{,}8 = 21{,}5W \quad \textbf{(4.127)}$$

Losses in the inductor:

$$P_{inductor} = P_{core} + P_W = 21{,}5 + 16{,}3 = 37{,}8W \quad \textbf{(4.128)}$$

Capacitor

The capacitor will be a film capacitor for the output filter

Type	Manufacturer/Reference	EPCOS MKP DC link
Aluminium Electrolytic	Capacitor value (µF)	3
	Voltage rating (V)	800
	ESR (mΩ)	7
	ESL (nH)	25
	$I_{rms\ @\ 20kHz}$ (A)	4,5

Losses in the output filter capacitor

$$P_{losses} = I_c{}^2 * (ESR) = 1{,}3^2 * (0{,}007) = 0{,}012W \quad \textbf{(4.129)}$$

12) Efficiency of the on board charger

Efficiency in the converter:

$$\eta = \frac{P_{out}}{P_{in}} = \frac{P_{in} - P_{losses}}{P_{in}} \quad \textbf{(4.130)}$$

With P_{losses} which represents the losses in the semiconductors, inductors, capacitors and transformer.

The efficiency of the on board charger is determined thanks to a Matlab script. This efficiency is calculated for an input current going from 6 to 16 Amps. The script takes care of the two control methods which were investigated before (see: Current loop and PWM). The way to determine the losses in the semiconductors, inductors, capacitors and transformer was describe previously. It is these expressions that can be find in the Matlab script. This script can be seen in Appendix XX with an explanation of how the data of the components, which allowed calculating the losses, was found. The script allows showing several things, the losses in the semi conductors, in the inductors, capacitors and transformer. The losses can also be put together, for example, losses in the power factor corrector, in the full bridge converter or in the whole on-board charger. The main results are shown below.

Losses in the power factor corrector components

The Figure 53 shows the losses in the rectifier bridge, Figure 54 the losses in the boost transistor, Figure 55 the losses in the boost diode, Figure 56 the losses in the boost inductor, Figure 57 the losses in the boost capacitor. The losses in the whole PFC can be viewed in Figure 58.

Losses in the input rectifier bridge

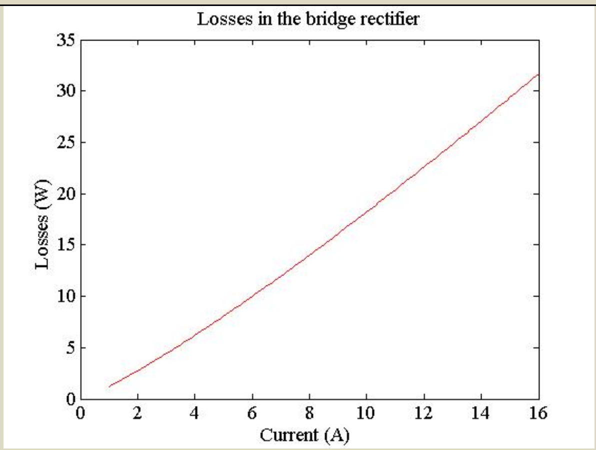


Figure 53 Losses in the rectifier bridge

Losses in the boost transistor

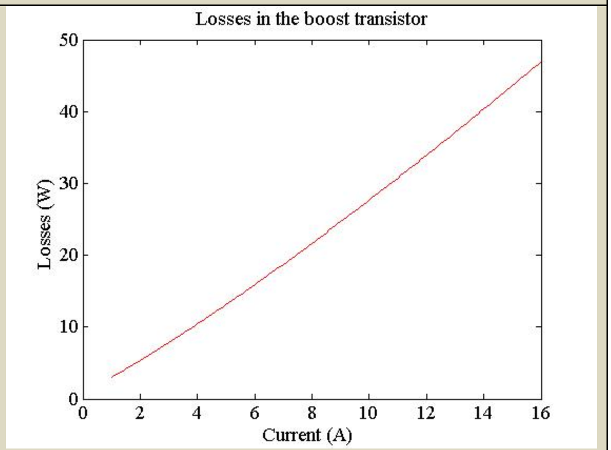


Figure 54 Losses in the boost transistor

Losses in the boost diode

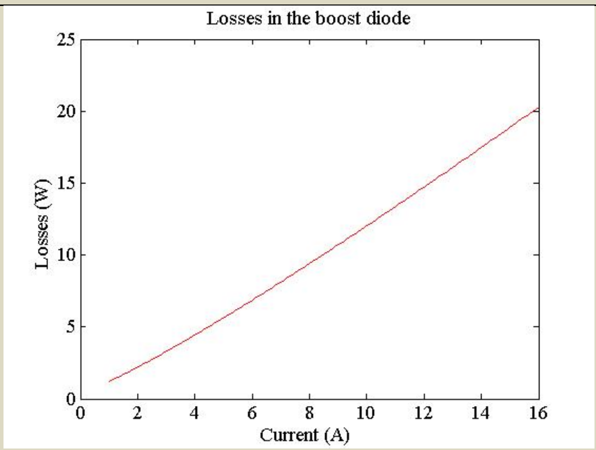


Figure 55 Losses in the boost diode

Losses in the boost inductor

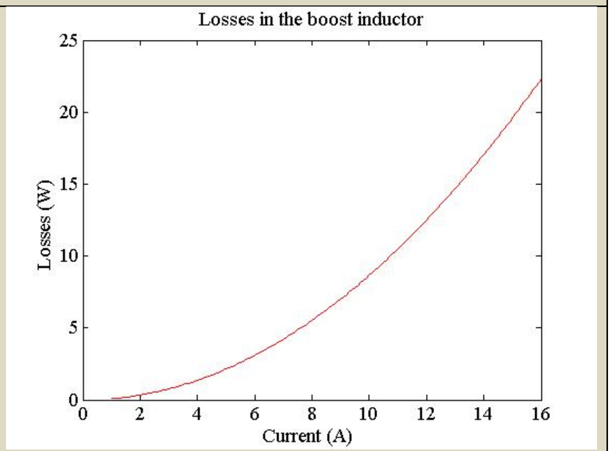


Figure 56 Losses in the inductor

Losses in the DC link capacitor

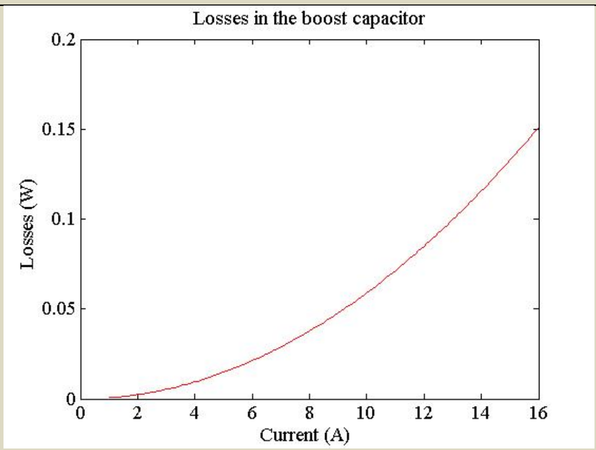


Figure 57 Losses in the capacitor

Losses in the Power Factor Corrector

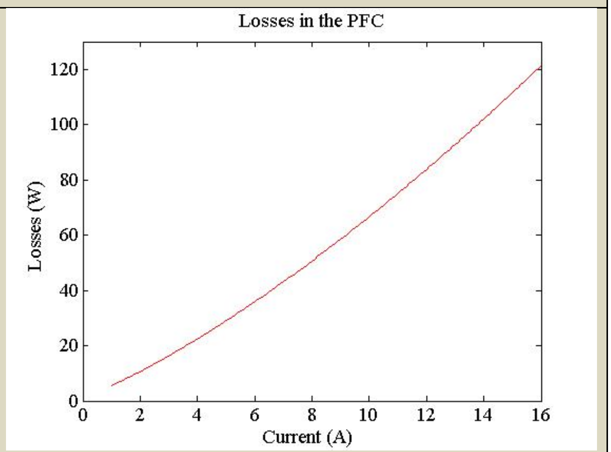


Figure 58 Losses in the PFC

The losses in the capacitor can be ignored, they are very small. The losses in the power factor corrector are not linear, that is the causes of some losses which are function of the square of the current.

Losses in the full bridge converter components with a 400V battery with phase shift control

The Figure 59 shows the losses in the inverter bridge, Figure 60 the losses in the transformer, Figure 61 the losses in the rectifier bridge, Figure 62 the losses in the inductor, Figure 63 the losses in the capacitor. The losses in the full bridge converter can be viewed Figure 64.

Losses in inverter	Losses in transformer

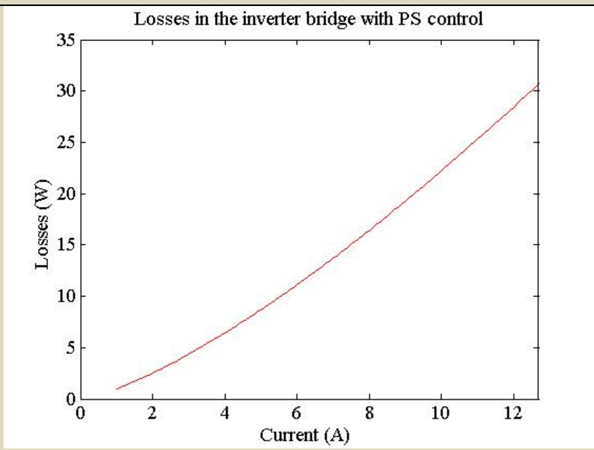
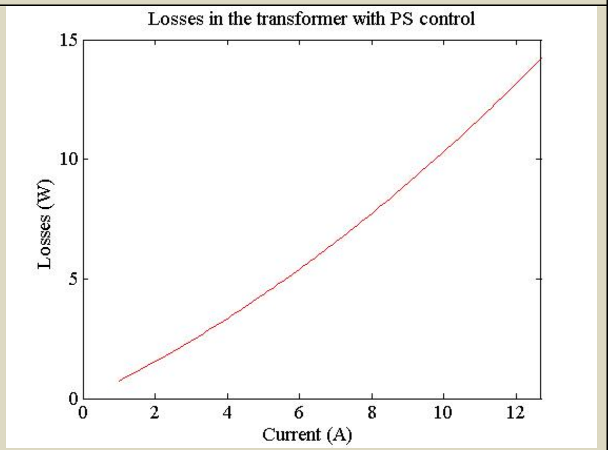

	

Figure 59 Losses in the inverter	Figure 60 Losses in the transformer
Losses in output rectifier bridge	**Losses in the filter inductor**

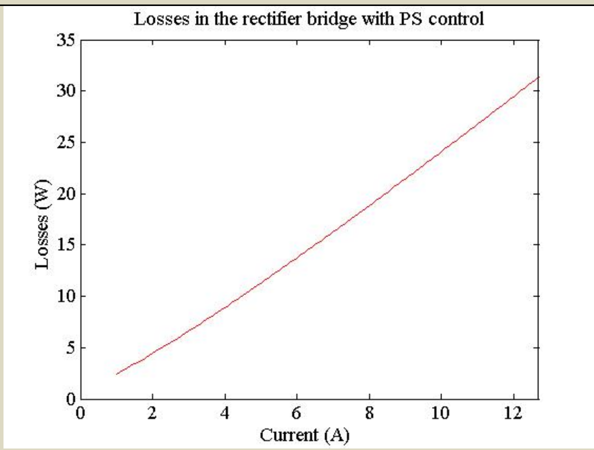
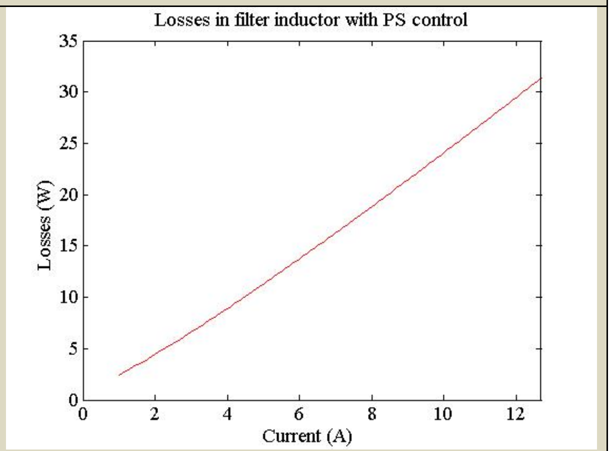

	

Figure 61 Losses in the rectifier bridge	Figure 62 Losses in the inductor
Losses in the filter capacitor	**Losses in the Full Bridge Converter**

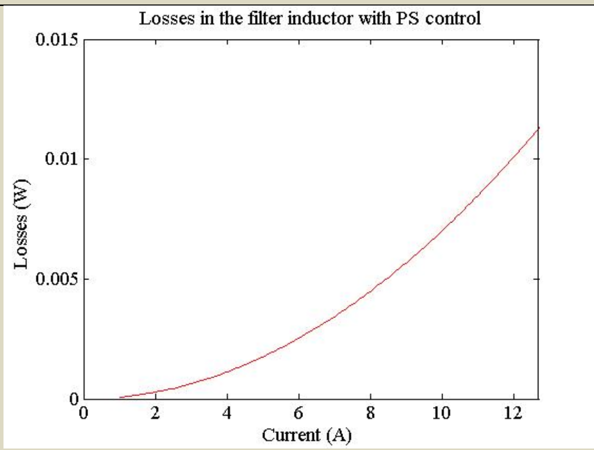
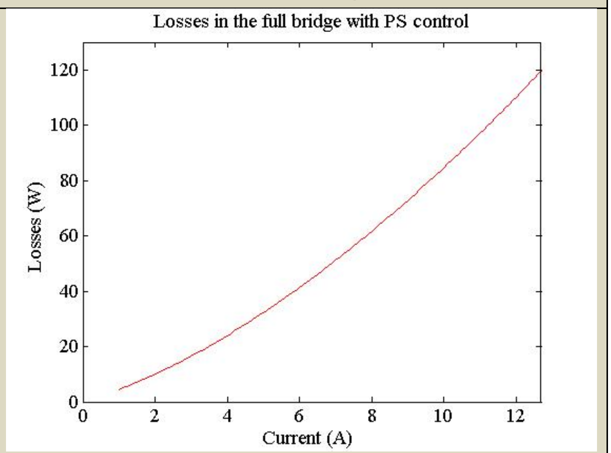

	

Figure 63 Losses in the capacitor	Figure 64 Losses in the full bridge converter

In the full bridge converter, the filter capacitor losses can be ignored too. The losses in the filter inductor are too high (about 41% of the full bridge converter losses are due to the inductor); it could be necessary to design another filter inductor. The losses in the full bridge converter are not linear, that is the cause of some losses which are a function of the square of the current.

Losses in the full bridge converter components with a 400V battery with duty cycle control

The Figure 65 shows the losses in the inverter bridge, Figure 66 the losses in the transformer, Figure 67 the losses in the rectifier bridge, Figure 68 the losses in the inductor, Figure 69 the losses in the capacitor. The losses in the full bridge converter can be viewed Figure 70.

Losses in inverter

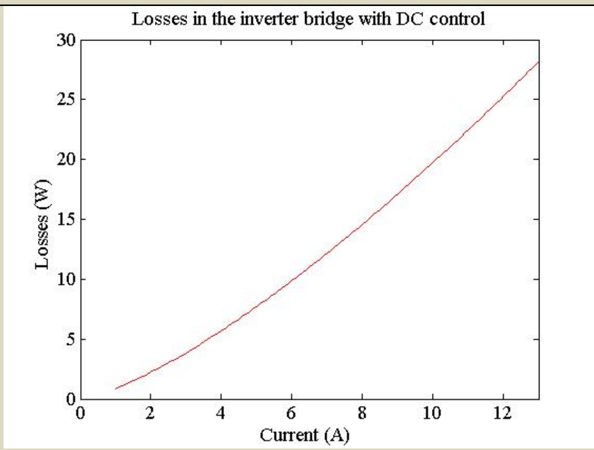


Figure 65 Losses in the inverter

Losses in transformer

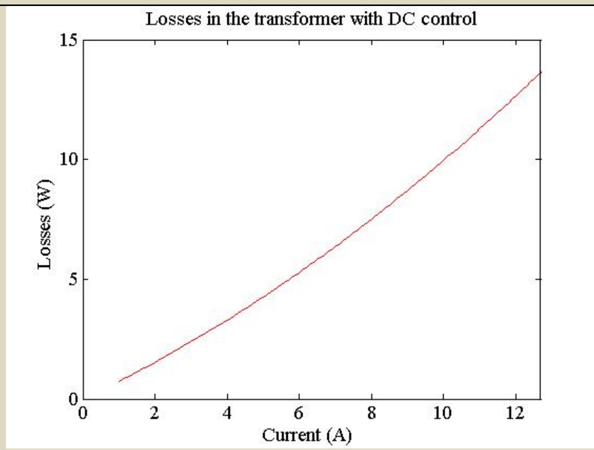


Figure 66 Losses in the transformer

Losses in output rectifier bridge

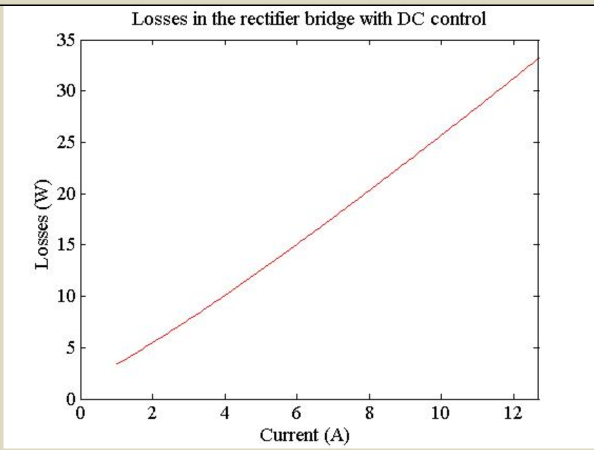


Figure 67 Losses in the rectifier bridge

Losses in the filter inductor

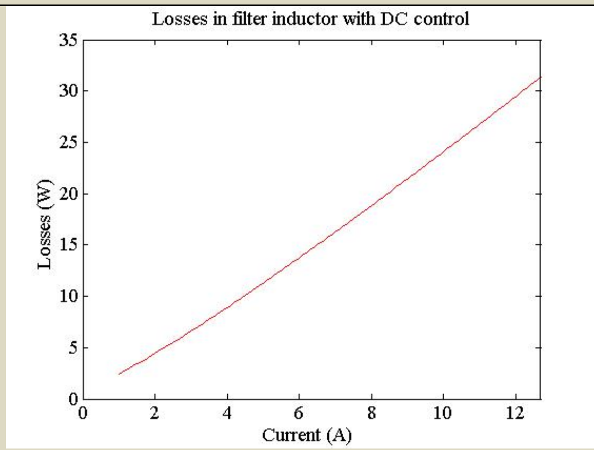


Figure 68 Losses in the inductor

Losses in the filter capacitor

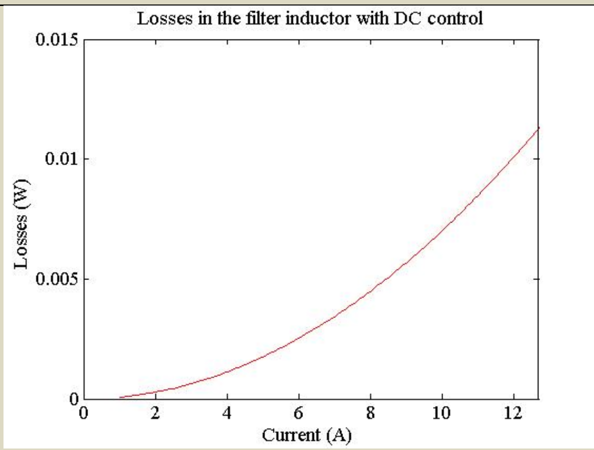


Figure 69 Losses in the capacitor

Losses in the Full Bridge Converter

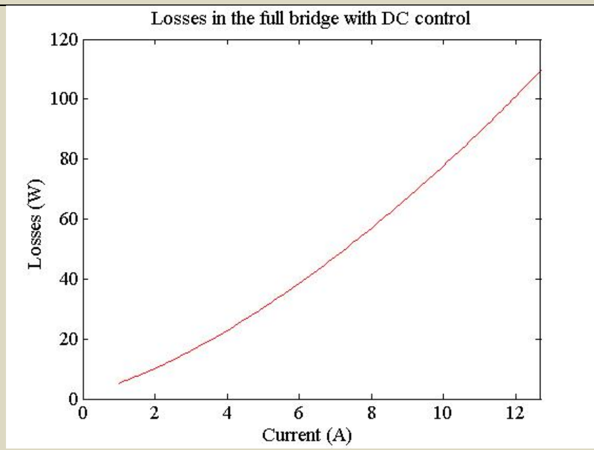


Figure 70 Losses in the full bridge converter

In the full bridge converter, the filter capacitor losses can be ignored too. The losses in the filter inductor are too high (about 42% of the full bridge converter losses are due to the inductor); it could be necessary to design another filter inductor. The losses in the full bridge converter are not linear, that is the cause of some losses which are a function of the square of the current.

Comparison of the efficiency in the on board charger function of the method control and the battery voltage

The efficiency of the OBC with a 300V battery can be viewed Figure 71. The Figure 72 shows the efficiency with a 400V battery.

Efficiency of the OBC with a 300V battery function of the control method

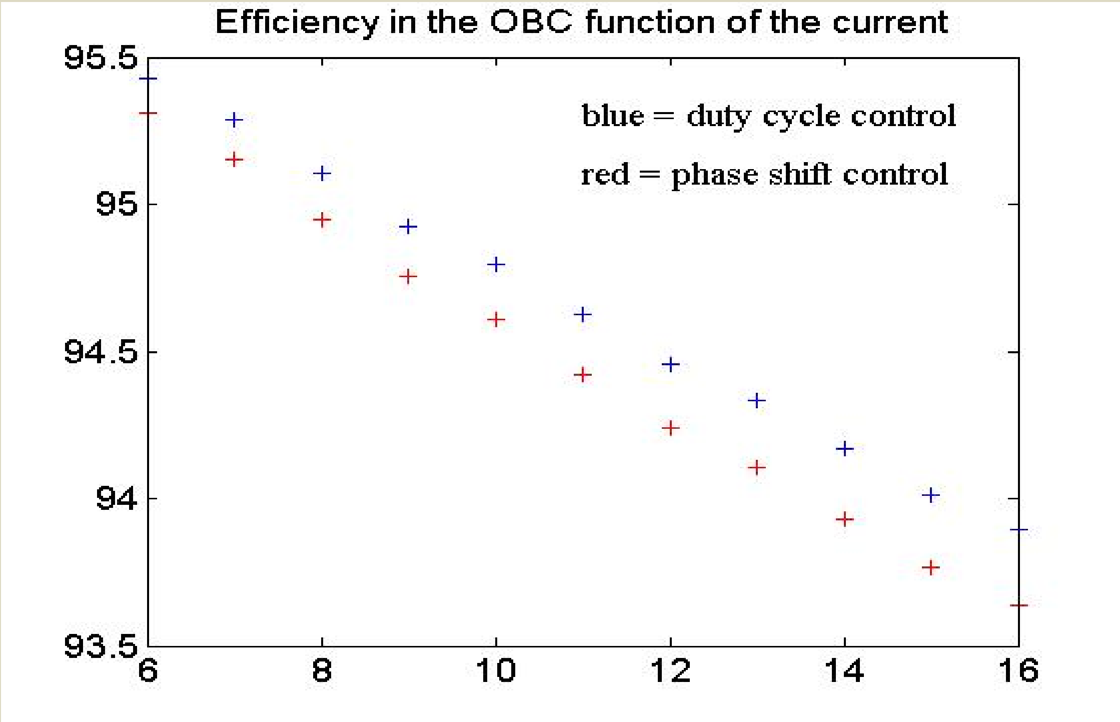


Figure 71 Efficiency of the OBC function of the control method at 300V

Efficiency of the OBC with a 400V battery function of the control method

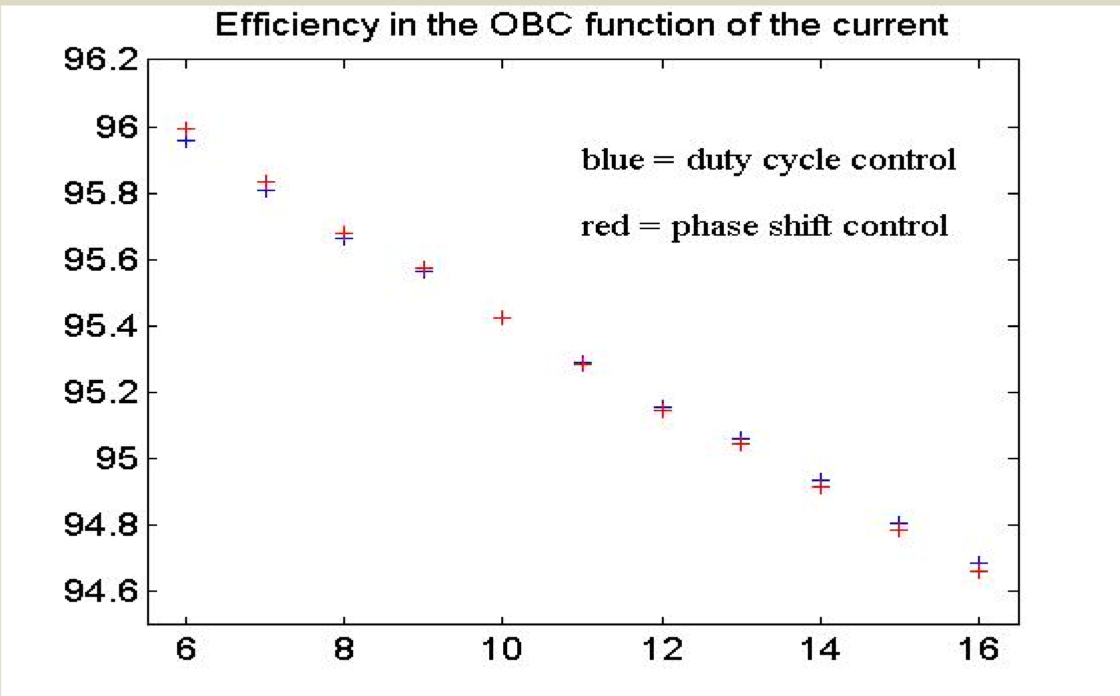


Figure 72 Efficiency of the OBC function of the control method at 400V

With a 400V battery, the efficiency is about the same whatever the control method. When the voltage is lower, the efficiency with the duty cycle control is better than with the phase shift control and, the lower the battery voltage is the higher the difference in efficiency is. Whatever the kind of control method, the higher the battery voltage is and the higher the efficiency is. The efficiency of the on-board charger is varying according to the battery voltage. That means that during the load, the efficiency is varying because the battery voltage is increasing.

The Table 4 shows the efficiency for a battery of 400V with a grid of 10 Amps and a grid of 16 Amps with the two different control methods. The following hypothesis can be made: during the load, the battery voltage increases with a linear progression.

	Efficiency (%)	Duty cycle control	Phase shift control
Input current $I_{in} = 16A$	$\eta_{U_{disloaded}}$	93,63	93,37
	$\eta_{U_{loaded}}$	94,56	94,53
	$\eta_{average}$	94,095	93,95
Input current $I_{in} = 10A$	$\eta_{U_{disloaded}}$	94,67	94,48
	$\eta_{U_{loaded}}$	95,37	95,37
	$\eta_{average}$	95,02	94,93

Table 4 Efficiency for a 400V battery

This table shows that, for a battery of 400V and whatever the input current, the duty cycle control has better efficiency. Moreover the lower the voltage of the battery is, the lower the losses in the full bridge converter are with the duty cycle control in comparison with the phase shift control, the duty cycle control is the method having the better efficiency.

Comparison between the unidirectional and the bidirectional OBC:

For this comparison, the phase shift control method is chosen. This method is probably the best one for the bidirectional OBC because the switching losses in each transistor can be ignored. The input inductor of the bidirectional charger is identical to the PFC inductor from the unidirectional OBC (current and inductor value) so the losses are roughly the same. The output filter is the same for both converters. The efficiency of the unidirectional charger was presented before. Figure 73 shows the efficiency in both converters with the phase shift control.

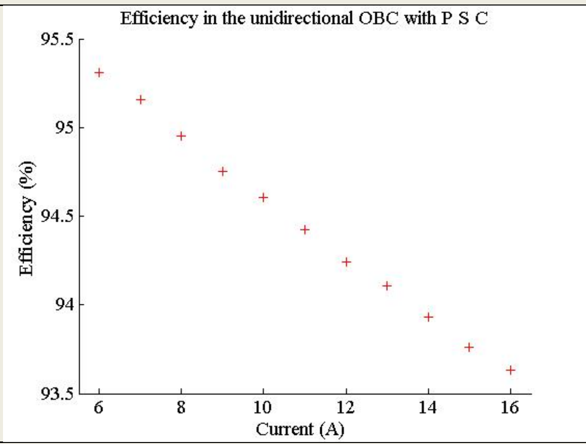
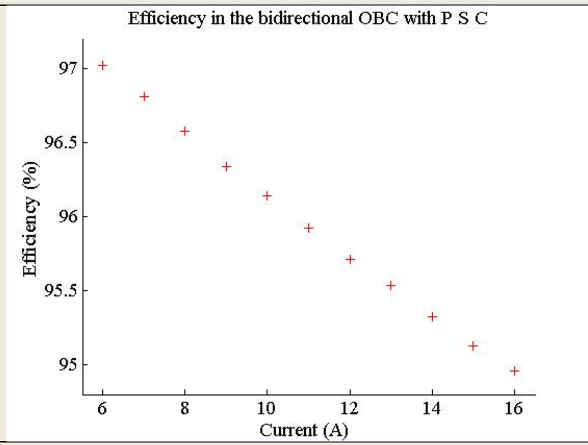
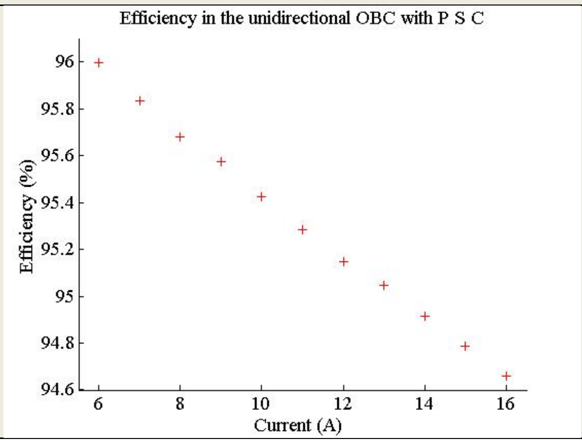
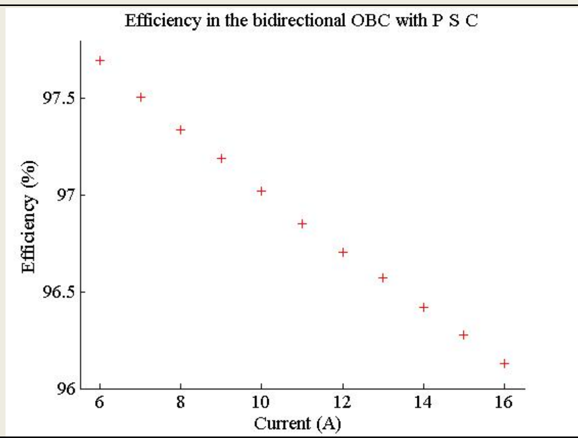
Phase Shift Control	
Unidirectional	Bidirectional
@ 300V	@ 300V
	
@ 400V	@ 400V
	

Figure 73 Efficiency comparison between bidirectional and unidirectional OBC

This Figure 73 shows that the bidirectional on board charger has a better efficiency than the unidirectional charger.

Table 5 shows the efficiency with a 10 Amps grid and a 16 Amps grid.

	Efficiency (%)	Unidirectional OBC	Bidirectional OBC
Input current $I_{in} = 16A$	$\eta_{U_{disloaded}}$	93,37	94,95
	$\eta_{U_{loaded}}$	94,53	96,13
	$\eta_{average}$	93,95	95,54
Input current $I_{in} = 10A$	$\eta_{U_{disloaded}}$	94,48	96,14
	$\eta_{U_{loaded}}$	95,37	97,02
	$\eta_{average}$	94,93	96,5

Table 5 Efficiency at 10 Amps and 16 Amps

The bidirectional on board charger has a better efficiency than the unidirectional OBC.

13) Refreshment of the semi conductor

Knowing all the losses in the semiconductors (switching and conduction losses), the following step is to design the heatsinks allowing to evacuate the heat. The maximum heat for the silicon is 150°C but, for the calculations, 125°C is taken, this value is the silicium junction temperature T_j. The ambient temperature T_a is 40°C. Figure 74 shows the thermal resistances in the system; $R_{TH\ JC}$ is the thermal resistance between the semi conductor junction and the case, $R_{TH\ CS}$ is the thermal resistance between the case and the heatsink and $R_{TH\ SA}$ is the thermal resistance between the heatsink and the ambient temperature. One semiconductor switch is associated to one heatsink. The objective is to find a suitable thermal resistance of the heatsink.

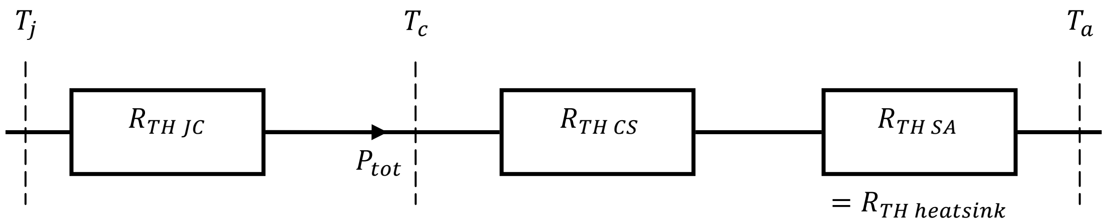


Figure 74 Thermal resistance in a system semiconductor/sink

R_{TH} is the total thermal resistance and **(4.131)**

$$R_{TH} = R_{TH\ JC} + R_{TH\ CS} + R_{TH\ SA}$$

Table 6 gives the power losses in the on-board charger semiconductor and theirs thermal resistances:

Semiconductor	Power losses (W)	Thermal res. $R_{jc}\ K/W$	Thermal res. $R_{cs}\ K/W$
Input bridge rectifier diode	16	1,2	0,5
Boost transistor	48,4	0,8	0,5
Boost diode	20,4	0,9	0,5
Inverter transistor	16	0,8	0,5
Output bridge rectifier diode	20,4	0,9	0,5

Table 6 Losses and thermal resistance of the OBC semiconductor

The use of a fan to cool the converter improves the dissipation power of the heatsinks. When the air speed is higher than 5m/s, the thermal resistance is four times lower.

Determination of the thermal resistance of the heatsink for the input bridge rectifier diode

$$P_{tot} = \frac{(T_j - T_a)}{R_{TH}} \quad \textbf{(4.132)}$$

Thermal resistance between the junction of the semiconductor and the ambient environment:

$$R_{TH} = \frac{(T_j - T_a)}{P_{tot}} = \frac{(125 - 40)}{16} = 5{,}25\ K/W \quad \textbf{(4.133)}$$

Thermal resistance of the heatsink for one diode:

$$R_{TH\,SA} = R_{TH} - \left(R_{TH\,JC} + R_{TH\,CS}\right) = 5{,}25 - (1{,}2 + 0{,}5) = 3{,}55\ K/W \tag{4.134}$$

It is possible to fix several diodes on one heatsink, the choice is to put 2 diodes or 4 diodes on the same heatsink for the input rectifier bridge diodes. This has the advantages to reduce the total volume of the heatsinks and to reduce the size of the on-board charger.

Thermal resistance of the heatsink with 2 or 4 diodes of the same heatsink:

$$R_{TH\,SA\,2\,diodes} = \frac{R_{TH\,SA}}{2} = \frac{3{,}55}{2} = 1{,}78\ K/W \tag{4.135}$$

$$R_{TH\,SA\,4\,diodes} = \frac{R_{TH\,SA}}{4} = \frac{3{,}55}{4} = 0{,}89\ K/W \tag{4.136}$$

With the use of a fan, the thermal resistance change with a factor of 4 so the thermal resistance of the heatsink:

$$R_{TH\,SA\,with\,FAN} = 4 * R_{TH\,SA} = 14{,}2\ K/W \tag{4.137}$$

Thermal resistance of the heatsink with 4 diodes of the same heatsink with fan:

$$R_{TH\,SA\,4\,diodes\,with\,pan} = \frac{R_{TH\,SA}}{4} = \frac{14{,}2}{4} = 3{,}55\ K/W \tag{4.138}$$

Determination of the thermal resistance of the heatsink for the boost transistor

The thermal resistance between the junction of the semiconductor and the ambient environment:

$$R_{TH} = \frac{(T_j - T_a)}{P_{tot}} = \frac{(125 - 40)}{48{,}4} = 1{,}76\ K/W \tag{4.139}$$

The thermal resistance of the heatsink for the boost transistor can be found as:

$$R_{TH\,SA} = R_{TH} - \left(R_{TH\,JC} + R_{TH\,CS}\right) = 1{,}76 - (0{,}8 + 0{,}5) = 0{,}46\ K/W \tag{4.140}$$

The thermal resistance of the heatsink for the boost transistor with a fan:

$$R_{TH\,SA\,with\,FAN} = 4 * R_{TH\,SA} = 1{,}84\ K/W \tag{4.141}$$

Determination of the thermal resistance of the heatsink for the boost diode

The thermal resistance between the junction of the semiconductor and the ambient environment:

$$R_{TH} = \frac{(T_j - T_a)}{P_{tot}} = \frac{(125 - 40)}{20{,}4} = 4{,}17\ K/W \tag{4.142}$$

The thermal resistance of the heatsink for the boost diode can be found as:

$$R_{TH\,SA} = R_{TH} - \left(R_{TH\,JC} + R_{TH\,CS}\right) = 4{,}17 - (0{,}9 + 0{,}5) = 2{,}77\ K/W \quad \textbf{(4.143)}$$

The thermal resistance of the heatsink for the boost diode with a fan:

$$R_{TH\,SA\,with\,FAN} = 4 * R_{TH\,SA} = 11\ K/W \quad \textbf{(4.144)}$$

The thermal resistance required to fix the boost diode and the boost transistor on the same heatsink

$$R_{TH\,without\,pan} = \frac{R_{TH\,SA\,diode} * R_{TH\,SA\,transistor}}{R_{TH\,SA\,diode} + R_{TH\,SA\,transistor}} = \frac{2{,}77 * 0{,}46}{2{,}77 + 0{,}46} = 0{,}40\ K/W \quad \textbf{(4.145)}$$

The thermal resistance required to fix the boost diode and transistor on the same heatsink with a fan:

$$R_{TH\,with\,pan} = \frac{R_{TH\,SA\,diode} * R_{TH\,SA\,transistor}}{R_{TH\,SA\,diode} + R_{TH\,SA\,transistor}} = \frac{11 * 1{,}84}{11 + 1{,}84} = 1{,}57\ K/W \quad \textbf{(4.146)}$$

Determination of the thermal resistance of the heatsink for the inverter transistor

The thermal resistance between the junction of the semiconductor and the ambient environment:

$$R_{TH} = \frac{(T_j - T_a)}{P_{tot}} = \frac{(125 - 40)}{16} = 5{,}31\ K/W \quad \textbf{(4.147)}$$

The thermal resistance of the heatsink for one transistor can be found as r:

$$R_{TH\,SA} = R_{TH} - \left(R_{TH\,JC} + R_{TH\,CS}\right) = 5{,}31 - (0{,}8 + 0{,}5) = 4{,}01\ K/W \quad \textbf{(4.148)}$$

The thermal resistance of the heatsink with 4 transistors of the same heatsink:

$$R_{TH\,SA\,4\,transistors} = \frac{R_{TH\,SA}}{4} = \frac{4{,}01}{4} = 1\ K/W \quad \textbf{(4.149)}$$

The thermal resistance of the heatsink for one transistor with a fan:

$$R_{TH\,SA\,with\,FAN} = 4 * R_{TH\,SA} = 16\ K/W \quad \textbf{(4.150)}$$

The thermal resistance of the heatsink with 4 transistors of the same heatsink with a fan:

$$R_{TH\,SA\,with\,FAN\,4\,transistors} = \frac{R_{TH\,SA\,with\,FAN}}{4} = \frac{16}{4} = 4\ K/W \quad \textbf{(4.151)}$$

Determination of the thermal resistance of the heatsink for the output bridge rectifier diode

The thermal resistance between the junction of the semiconductor and the ambient environment:

$$R_{TH} = \frac{(T_j - T_a)}{P_{tot}} = \frac{(125 - 40)}{20{,}4} = 4{,}17\ K/W \quad \textbf{(4.152)}$$

The thermal resistance of the heatsink for one diode can be found as:

$$R_{TH\,SA} = R_{TH} - \left(R_{TH\,JC} + R_{TH\,CS}\right) = 4{,}17 - (0{,}9 + 0{,}5) = 2{,}77\ K/W \quad \textbf{(4.153)}$$

The thermal resistance of the heatsink with 2 or 4 diodes of the same heatsink:

$$R_{TH\,SA\,2\,diodes} = \frac{R_{TH\,SA}}{2} = \frac{2{,}77}{2} = 1{,}38\ K/W \quad \textbf{(4.154)}$$

$$R_{TH\,SA\,4\,diodes} = \frac{R_{TH\,SA}}{4} = \frac{2{,}77}{4} = 0{,}69\ K/W \quad \textbf{(4.155)}$$

The thermal resistance of the heatsink for one diode with a fan:

$$R_{TH\,SA\,with\,FAN} = 4 * R_{TH\,SA} = 11\ K/W \quad \textbf{(4.156)}$$

The thermal resistance of the heatsink with 4 diodes of the same heatsink with a fan:

$$R_{TH\,SA\,4\,diodes\,with\,fan} = \frac{R_{TH\,SA\,with\,FAN}}{4} = \frac{11}{4} = 2{,}75\ K/W \quad \textbf{(4.157)}$$

Choice of the heatsinks

The choice of the heatsinks is a compromise between the volume of the heatsink and the price.

On-board charger without pan

The Table 7 gives the thermal resistances and power losses in the on-board charger without pan.

	$R_{TH\,SA}$ (K/W)	P_{dissip} (W)	$R_{TH\,SA\,2D}$ (K/W)	$P_{dissip\,2D}$ (W)	$R_{TH\,SA\,4D}$ (K/W)	$P_{dissip\,4D}$ (W)
Input bridge rectifier diode	3,55	16	1,78	32	0,89	64
Boost transistor	0,46	48,4				
Boost diode	2,77	20,4				
Boost diode and transistor	0,40	68,8				
Inverter transistor	4,01	16	2	32	1	64
Output bridge rectifier diode	2,77	20,4	1,38	40,8	0,69	81,6

Table 7 Thermal resistances and power losses in the OBC components without pan

Input bridge rectifier diode

	Manufacturer / Reference	Thermal resistance (K/W)	Maximum power (W)	Volume (cm^3)	Weight (g)	Price (kr)
1 diode	Austerlitz electronic KS65-75E	3	20	117	84	70
2 diodes	Austerlitz electronic KS105-75E	1,8	50	150	135	128
4 diodes	Austerlitz electronic KS160-100E	0,65	75	640	1000	247

The choice of the heatsinks for the diodes in the input bridge rectifier is the Austerlitz electronic KS105-75E. This one is the cheapest, the lightest and the one which take the least volume.

Boost transistor and diode

	Manufacturer / Reference	Thermal resistance (K/W)	Maximum power (W)	Volume (cm^3)	Weight (g)	Price (kr)
transistor	Austerlitz electronic KS216-100E	0,4	100	1793	2040	533
diode	Austerlitz electronic KS88-75E	2	30	231	215	85
Transistor and diode	Austerlitz electronic KS216-100E	0,4	100	1793	2040	533

The heatsink chooses for the boost transistor and diode is the Austerlitz electronic KS216-100E. The diode and the transistor are fixed on this heatsink.

Inverter transistor

	Manufacturer / Reference	Thermal resistance (K/W)	Maximum power (W)	Volume (cm^3)	Weight (g)	Price (kr)
1 transistor	Austerlitz electronic KS65-37,5E	4	20	58,5	42	70
2 transistors	Austerlitz electronic KS105-75E	1,8	50	150	135	128
4 transistors	Austerlitz electronic KS160-100E	0,65	75	640	1000	247

For the inverter transistor, the best choice is to take 4 heatsinks Austerlitz electronic KS65-37,5E. It's the most expensive choice but it's also the lightest and the solution using the least volume.

Output bridge rectifier diode

	Manufacturer / Reference	Thermal resistance (K/W)	Maximum power (W)	Volume (cm^3)	Weight (g)	Price (kr)
1 diode	Austerlitz electronic KS88-75E	2	30	231	215	61
2 diodes	Austerlitz electronic KS105-100E	1,8	75	200	180	116
4 diodes	Austerlitz electronic K216-75E	0,60	100	1345	1530	483

For the output rectifier bridge diode, the best choice is to take 2 heatsinks Austerlitz electronic KS105-100E. It's the cheapest choice, the lightest and the solution using the least volume.

On-board charger with pan

The Table 8 gives the thermal resistances and power losses in the on-board charger with pan.

	$R_{TH\ SA}$ (K/W)	P_{dissip} (W)	$R_{TH\ SA\ 2D}$ (K/W)	$P_{dissip\ 2D}$ (W)	$R_{TH\ SA\ 4D}$ (K/W)	$P_{dissip\ 4D}$ (W)
Input bridge rectifier diode	14,2	16	7,1	32	3,55	64
Boost transistor	1,84	48,4				
Boost diode	11	20,4				
Boost diode and transistor	1,57	68,8				
Inverter transistor	16	16	8	32	4	64
Output bridge rectifier diode	11	20,4	5,5	40,8	2,75	81,6

Table 8 Thermal resistances and power losses in the OBC with pan

Input bridge rectifier diode

	Manufacturer / Reference	Thermal resistance (K/W)	Maximum power (W)	Volume (cm^3)	Weight (g)	Price (kr)
1 diode	Austerlitz electronic KS57.2-50E	5	20	72	65	74
2 diodes	Austerlitz electronic KS105-75E	1,8	50	150	135	128
4 diodes	Austerlitz electronic KS105-100E	1,3	75	200	180	115

The choice of the heatsinks for the diodes in the input bridge rectifier is the Austerlitz electronic KS105-100E. This one is the cheapest, the lightest and the one which take the least volume.

Boost transistor and diode

	Manufacturer / Reference	Thermal resistance (K/W)	Maximum power (W)	Volume (cm^3)	Weight (g)	Price (kr)
transistor	Austerlitz electronic KS105-75E	1,8	50	150	135	128
diode	Austerlitz electronic KS94-100E	2,9	20	141	168	51
Transistor and diode	Austerlitz electronic KS105-100E	1,3	75	200	180	115

The heatsink chosen for the boost transistor and diode is the Austerlitz electronic KS105-100E. The diode and the transistor are fixed on this heatsink. This solution is the cheapest, the lightest and the one which take the least volume.

Inverter transistor

	Manufacturer / Reference	Thermal resistance (K/W)	Maximum power (W)	Volume (cm^3)	Weight (g)	Price (kr)
1 transistor	Austerlitz electronic KS65-37,5E	4	20	58,5	42	70
2 transistors	Austerlitz electronic KS105-75E	1,8	50	150	135	128
4 transistors	Austerlitz electronic KS105-100E	1,3	75	200	180	115

For the inverter transistor, the best choice is to take the heatsinks Austerlitz electronic KS105-100E. It's the least expensive choice, the solution using the least volume but it's not the lightest.

Output bridge rectifier diode

	Manufacturer / Reference	Thermal resistance (K/W)	Maximum power (W)	Volume (cm^3)	Weight (g)	Price (kr)
1 diode	Austerlitz electronic KS88-75E	2	30	231	215	61
2 diodes	Austerlitz electronic KS105-100E	1,8	75	200	180	116
4 diodes	Austerlitz electronic K216-75E	0,60	100	1345	1530	483

For the output rectifier bridge diode, the best choice is to take 2 heatsinks Austerlitz electronic KS105-100E. It's the cheapest choice, the lightest and the solution using the least volume.

Comparison between the two solutions:

$$Weight_{heatsink\ without\ fan} = 2 * 135 + 2040 + 4 * 42 + 2 * 180 = 2{,}839\ kg \tag{4.158}$$

$$Weight_{heatsink\ with\ fan} = 180 + 180 + 180 + 2 * 180 = 0{,}900\ kg \tag{4.159}$$

$$Volume_{heatsink\ without\ fan} = 3000 + 1793 + 4 * 58{,}5 + 400 = 2727\ cm^3 \tag{4.160}$$

$$Volume_{heatsink\ with\ fan} = 200 + 200 + 200 + 2 * 200 = 1000\ cm^3 = 1\ dm^3 \tag{4.161}$$

$$Cost_{heatsink\ without\ fan} = 2 * 128 + 533 + 4 * 70 + 2 * 115 = 1300\ kr \tag{4.162}$$

$$Cost_{heatsink\ with\ fan} = 115 + 115 + 115 + 2 * 115 = 575\ kr \tag{4.163}$$

	OBC without fan	OBC with fan	Reduction ratio
Weight (kg)	2,839	0,9	68,3%
Volume (dm^3)	2,73	1	63,4%
Cost (kr)	1300	575	55,8%

Table 9 Comparison of the heatsinks with or without pan

The Table 9 shows the interest of a fan in this on board charger. A fan allows the reduction of 68% of the weight and a reduction of 63% of the volume of the heatsinks in the converters. In a nutshell, the on board charger is really smaller and lighter.

14) Weight, volume and price of the main components of the on-board charger

Weight

The weight of the main components can be determined. The components used to calculate this weight are the heatsinks, the transformer, the inductor and the DC link capacitor.

$$Weight_{comp} = Weight_{capa} + Weight_{ind} + Weight_{transf} + Weight_{sink}$$
$$= 0{,}44 + 0{,}497 + 0{,}497 + 0{,}52 + 0{,}9 = 2{,}85kg$$

We can assume that these components represent about 60% of the whole on-board charger.

$$Weight_{OBC} = \frac{Weight_{comp}}{0{,}6} = \frac{2{,}85}{0{,}6} = 4{,}75\ kg$$

The weight of the on-board charger is around of 4,75 kg.

Volume

The volume of the main components can be determined. The components used to calculate this volume are the same using to determine the previously weight.

$$Volume_{comp} = Volume_{capa} + Volume_{ind} + Volume_{transf} + Volume_{sink}$$

$$= 0{,}34 + 0{,}071 + 0{,}071 + 0{,}083 + 1 = 1{,}57dm^3$$

We can assume that these components represent about 60% of the whole on-board charger.

$$Volume_{OBC} = \frac{Volume_{comp}}{0{,}6} = \frac{1{,}57}{0{,}6} = 2{,}7\ dm^3$$

The volume of the on board charger should be around 2,7 dm^3. This volume can be contained, for example, in a box of the following dimensions:

Length = 22,5cm, Width = 12cm, Height=10cm

Price

The price of the main components can be determined; these components are the one of the power part, diodes, transistors, inductor cores, transformer core, capacitors and heatsinks.

$$Price_{diodes} = 8 * 5{,}2 + 5 * 2{,}2 = 52{,}6\ €$$

$$Price_{transistors} = 5 * 5{,}3 = 16{,}5\ €$$

$$Price_{inductors} = 2 * 13{,}3 = 26{,}6\ €$$

$$Price_{transformer} = 2 * 10{,}1 = 20{,}2\ €$$

$$Price_{capacitor} = 41 + 8 = 49\ €$$

$$Price_{heatsinks} = 57{,}5\ €$$

$$Price_{total} = 222{,}4\ €$$

The total price of the power part components of the on board charger is 222€. This price is enough high. Several things can reduce this price:

- The diode *RURP3060* are expensive (roughly 5,2€ per diode). This diode can be changed with another cheaper. A new diode can decrease the efficiency of the on-board charger so it's interesting to compare losses between the *RURP3060* and the new one.
- The prices given before are the price when only one component is bought. The prices for one hundred components reduce by two the previous price.

15) Time needed to load the battery with the on-board charger

Time to load a battery with 16A:

A PHEV battery size could have between 6 and 16kWh. In order to determine the charging time, a battery of 14kWh is used. It can be considered that the battery is empty when the State Of Charge level is 30% corresponding.

The efficiency of this system is 94,1% under 16A. That means that if the input current is 16A, the power fed into the battery will be:

$$P_{batt} = \eta * P_{in} = 0{,}941 * 3680 = 3463\ W \tag{4.164}$$

Battery fully charged:

$$W_{batt} = 14\ kWh$$

Battery disloaded:

$$W_{batt\ disload} = 0{,}3 * W_{batt} = 0{,}3 * 14\ kWh = 4{,}2\ kWh \tag{4.165}$$

Energy needed to load the battery:

$$W_{needed} = W_{batt\ load} - W_{batt\ disload} = 9{,}8kWh \tag{4.166}$$

Energy going to the battery during the load:

$$W_{load} = \eta * P_{in} * t = 0{,}941 * 3680 * 1 = 3463\ Wh \tag{4.167}$$

Time needed to load the battery:

$$t = \frac{W_{needed}}{W_{load}} = \frac{9800}{3463} = 2{,}83 \rightarrow 2\ hours\ and\ 50\ minutes \tag{4.168}$$

Time to load a battery with 10A:

The time needed with a 10 amps grid can be determined by the same way.

The efficiency of this system is 95% under 10A. That means that if the input current is 10A, the power fed into the battery will be

$$P_{batt} = \eta * P_{in} = 0{,}95 * 2300 = 2185\ W. \tag{4.169}$$

Battery fully charged:

$$W_{batt} = 14\ kWh$$

Battery disloaded:

$$W_{batt\ disload} = 0{,}3 * W_{batt} = 0{,}3 * 14\ kWh = 4{,}2\ kWh \tag{4.170}$$

Energy needed to load the battery:

$$W_{needed} = W_{batt\ load} - W_{batt\ disload} = 9{,}8kWh \tag{4.171}$$

Energy going to the battery during the load:

$$W_{load} = \eta * P_{in} * t = 0{,}95 * 2300 * 1 = 2185\ Wh \tag{4.172}$$

Time needed to load the battery:

$$t = \frac{W_{needed}}{W_{load}} = \frac{9800}{2185} = 4{,}49 \rightarrow 4\ hours\ and\ 29\ minutes \tag{4.173}$$

Time to load a battery with 16A for a distance of 40km:

An average drive distance is about 40km a day. A Plug-in Hybrid Electrical Vehicle (PHEV) uses about 2 kWh to cover 10kilometers.

The efficiency of this system is 94,1% under 16A.

$$P_{batt} = \eta * P_{in} = 0{,}941 * 3680 = 3463\ W \tag{4.174}$$

Battery fully charged:

$$W_{batt} = 14\ kWh$$

Energy needed to cover 40Km:

$$W_{40km} = 4 * 2\ kWh = 8\ kWh \tag{4.175}$$

Battery disloaded:

$$W_{batt\ disload\ after\ 40km} = W_{batt} - W_{40km} = 14 - 8 = 6\ kWh \tag{4.176}$$

Energy needed to load the battery:

$$W_{needed} = W_{40km} = 8\ kWh \tag{4.177}$$

Energy going to the battery during the load:

$$W_{load} = \eta * P_{in} * t = 0{,}95 * 3600 * 1 = 3463\ Wh \tag{4.178}$$

Time needed to load the battery:

$$t = \frac{W_{needed}}{W_{load}} = \frac{8000}{3463} = 2{,}31 \rightarrow\ 2\ hours\ and\ \ 19\ minutes \tag{4.179}$$

Time to load a battery with 10A for a distance of 40km:

The time needed with a 10 amps grid current to load the battery after a route of 40km can be determined by the same way.

Energy going to the battery during the load:

$$W_{load} = \eta * P_{in} * t = 0{,}95 * 2300 * 1 = 2185\ Wh \quad \textbf{(4.180)}$$

Time needed to load the battery:

$$t = \frac{W_{needed}}{W_{load}} = \frac{8000}{2185} = 3{,}66 \rightarrow 3\ hours\ and\ 40\ minutes \quad \textbf{(4.181)}$$

16) Cost to load a 14kWh battery with the on-board charger

Price to load the battery when it is fully disloaded:

The price of 1 kWh of electricity provided by the Swedish electricity company is about 1 SEK. If we considered that the change ratio between SEK and €, the price for 1 kWh is about 0,1€.

The energy needed to load the battery is $W_{needed} = 9{,}8kWh$.

The energy took on the grid is:

$$W_{toke} = \frac{W_{needed}}{\eta} = \frac{9{,}8}{0{,}941} = 10{,}41kWh \quad \textbf{(4.182)}$$

The price to load a battery fully disloaded is

$$Cost_{fully\ disload} = Cost_{\ 1kWh} * W_{toke} = 0{,}1 * 10{,}41 \quad \textbf{(4.183)}$$

$Cost_{fully\ disloaded} = 1{,}04$ €

The price to load a battery is about 1€

Price to load the battery for a 40km journey:

The price is still 0,1€ for 1 kWh.

The energy needed to load the battery is $W_{40km} = 8kWh$.

The energy took on the grid is

$$W_{toke} = \frac{W_{needed}}{\eta} = \frac{8}{0{,}941} = 8{,}5kWh \quad \textbf{(4.184)}$$

The price to load a battery after a 40km journey is

$$Cost_{40km} = Cost_{\ 1kWh} * W_{40km} = 0{,}1 * 8{,}5 \quad \textbf{(4.185)}$$

$Cost_{40km} = 0{,}85$ €

The price to load a battery is about 0,85€

Comparison between electrical cars and petrol cars:

In this part, the price for one year will be determined. On one hand, the price for a PHEV will be calculated, and on an other hand, the price for a petrol car.

This working hypothesis is setting:

The PHEV uses only electrical energy during the short journeys (lower than 50 kilometres)

During the year , 10 journeys with a distance of 500km
10 journeys with a distance of 150km
10 journeys with a distance of 100km
All the other days, a distance of 40km is covered.

The total distance covered during the year is assumed to be 20000 kilometres.

<u>Petrol cars:</u>

The price of petrol varying between 1 and 1,5€ a litre (depending on the price of the oil barrel). To determine the cost during one year, the price of the petrol is assumed to be 1,25€.

A medium car consumes about $C_{petrol} = 6$ litres of petrol for one hundred kilometres. To make the 20000 kilometres, a car needs

$$Q_{petrol} = Distance * Consumption = 20000 * \frac{6}{100} = 1200\ litres \qquad \textbf{(4.186)}$$

$$Cost_{petrol/year} = 1{,}25 * 1200 = 1500€ \qquad \textbf{(4.187)}$$

The cost, for one year of petrol, is about 1500€

<u>Plug-in Hybrid Electrical Vehicle:</u>

Several hypothesis are to be are laid down; the price of the petrol is still 1,25€ per litre. For the shorter journey (lower than 40km), the car is only propelled with electrical energy. For the other journey, car is propelled by internal combustion and electrical engines, petrol and electrical energies. A hybrid electrical vehicle reduces its petrol consumption by 25% compared to a car of same range (5).

Consumption with both energies:

$$Cost_{petrol/year} = 1{,}25 * 1200 = 1500€ \qquad \textbf{(4.188)}$$

$$C_{hybrid} = (1 - 0{,}25) * C_{petrol} = 0{,}75 * 6 = 4{,}5\ Litres/100km \quad \textbf{(4.190)}$$

Cost for the journey of 40km:

The cost of load was determined before for one journey of 40 kilometres $Cost_{40km} = 0{,}85$ €

In a year, we assumed that 330 journeys of this type are made.

$$Cost_{40km\ year} = 0{,}8\ € * 330 = 281€ \quad \textbf{(4.191)}$$

Cost of the other journey:

These journeys are 30 of them. For this journey, the cost is determined by the petrol consumption during the journey and the cost of the load of the battery.

These 30 journeys represent 7500 kilometres.

Cost of petrol for these journeys:

$$Q_{petrol} = Distance * Consumption = 7500 * \frac{4{,}5}{100} = 338\ litres \quad \textbf{(4.192)}$$

$$Cost_{petrol/year} = 1{,}25 * 338 = 422{,}5€ \quad \textbf{(4.193)}$$

Cost of electricity for these journeys:

At the end of the journey, the battery has to be fully loaded in order to be used for a classical journey he day after. That means that 30 loads have to be adding to the petrol cost to have the total cost for these journeys. The hypothesis is that at the end of the day, the State Of Charge is about 50%.

Battery fully charged:

$$W_{batt} = 14\ kWh$$

Battery disloaded:

$$W_{batt\ disload} = 0{,}5 * W_{batt} = 0{,}7 * 14\ kWh = 7\ kWh \quad \textbf{(4.194)}$$

Energy needed to load the battery:

$$W_{needed} = W_{batt\ load} - W_{batt\ disload} = 7\ kWh \quad \textbf{(4.195)}$$

The energy taken from the grid is:

$$W_{toke} = \frac{W_{needed}}{\eta} = \frac{7}{0{,}941} = 7{,}44kWh \quad \textbf{(4.196)}$$

The price to load a battery fully disloaded is:

$$Cost_{70\%} = Price_{\ 1kWh} * W_{needed} = 0{,}1 * 7{,}44 \quad \textbf{(4.197)}$$

$Cost_{70\%} = 0{,}75$ €

The 30 loads $Cost_{30\ loads}$ cost:

$$Cost_{30\ loads} = 30 * Cost_{70\%} = 30 * 0{,}7 = 22{,}5€ \quad \textbf{(4.198)}$$

The cost for one year of journeys with a hybrid vehicle is:

$$\begin{aligned} Cost_{hybrid/year} &= Cost_{30\ loads} + Cost_{petrol/year} + Cost_{40km/year} \\ &= 22{,}5 + 422{,}5 + 281 \end{aligned} \quad \textbf{(4.199)}$$

$Cost_{hybrid/year} = 726€$

The driving energy cost, for one year with a hybrid electrical vehicle, is about 726€. This cost is really lower than the one for a petrol car (1500€ a year). That means that the cost reduction with a PHEV is about 51% per year.

Forecasting on 10 years

The hypothesis is he same that previously.

Fuel car

The cost of the fuel for one year:

$$Cost_{petrol/year} = 1{,}25 * 1200 = 1500€$$

The cost of the fuel for ten years:

$$Cost_{petrol/10years} = 10 * Cost_{petrol/year} = 10 * 1500 = 15000€$$

The cost of the fuel for ten years is roughly 15000€.

Plug in hybrid electrical vehicle

The cost of the fuel and electricity for one year:

$$Cost_{hybrid/year} = 726€$$

The cost for ten years:

$$Cost_{hybrid/10years} = 10 * Cost_{hybrid/year} = 10 * 726€ = 7260€$$

The savings realized on ten years are 7740€.

V) Conclusion

In this master thesis, the theory of the on-board charger was investigated, including the power factor corrector, and the full bridge converter, power and control part. This converter was, then, simulated with Matlab Simulink. The simulation allowed the correction and the improvement of some problem. The components of the power part, the heatsinks were chosen and the magnetic components, snubbers and filter were designed. The losses in each component was calculated in order to determined the efficiency of the on board charger. The efficiency varies function of the control method and in this converter the duty cycle control has a better efficiency; the efficiency is 94,1% for a 400V battery with a 16Amps grid current. The approximate weight of the on-board charger is 3,2 kg and the volume is roughly 2,7 dm^3. If we considered that the length of life of the cars is ten years, the savings are roughly 7700€ with a PHEV.

The future works is to write a FPGA program which takes care of the method control of the full bridge converter and the end of load of the battery when the battery is totally loaded. Design the control part, with the choice of the FPGA, the supplies for control part, and the driver circuits. One good thing should improve the on-board charger in order to provide a very fast load during a short time (i.e. a 30 minutes load to drives 30kilometres).

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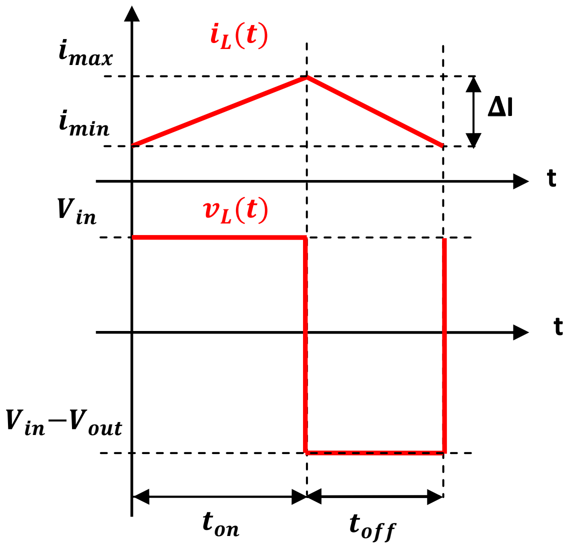
APPENDIX

1) PFC inductor value



$$t_{on} = \frac{L}{V_{in}}(I_{max} - I_{min})$$

$$t_{off} = \frac{L}{V_{out} - V_{in}}(I_{min} - I_{max})$$

$$f_{swi} = \frac{1}{t_{on} + t_{off}}$$

$$\Delta I = I_{max} - I_{min}$$

Figure 1 Current and voltage in the inductor

$$f_{swi} = \frac{V_{in}(V_{out} - V_{in})}{L\ \Delta I\ V_{out}}$$

$$V_{in} = V_M \sin(\omega t)$$

$$f_{swi} = \frac{V_M \sin(\omega t)\ (V_{out} - V_M \sin(\omega t)\ )}{L\ \Delta I\ V_{out}}$$

$\frac{df_{swi}}{dt} = 0$ gives the maximum value

The maximum value is for $\sin(\omega t) = \frac{V_{out}}{2V_M}$

$$f_{swi\,max} = \frac{V_M \frac{V_{out}}{2V_M}\ (V_{out} - \frac{V_{out}}{2V_M})}{L\ \Delta I\ V_{out}}$$

$$f_{swi\,max} = \frac{V_{out}}{4\ L\ \Delta I}$$

<u>The inductor value is:</u>

$$L = \frac{V_{out}}{4\ f_{swi\,max} \Delta I}$$

2) PFC capacitor value (DC link)

$v_{out}(t) = V_{out} + \delta v_{out}(t)$

Where V_{out} is the mean value of $v_{out}(t)$ and $\delta v_{out}(t)$ the 100Hz ripple of $v_{out}(t)$.

$$P_{grid}(t) = V_M I_M \, sin^2(\omega t) = \frac{V_M I_M}{2}[1 - \cos(2\omega t)]$$

$$P_{DC\ link\ (t)} = v_{out}(t).\, i_D(t) = v_{out}(t)\left(C\frac{dv_{out}(t)}{dt} + \frac{v_{out}(t)}{R}\right) = V_{out}\left(C\frac{d\delta v_{out}(t)}{dt} + \frac{V_{out}}{R}\right)$$

- Mean power: $\frac{V_M I_M}{2} = \frac{V_{out}{}^2}{R}$
- Ripple power: $-\frac{V_M I_M}{2}\cos(2\omega t) = C\ V_{out}\frac{d\delta v_{out}(t)}{dt}$

$v_{out}(t)$ ripple is $\delta v_{out}(t) = -\frac{V_M I_M}{4\,C\,\omega\,V_{out}}\sin(2\omega t)$

Peak to peak amplitude value is $\Delta V_{out} = \frac{V_M I_M}{2\,C\,\omega\,V_{out}}$

The capacitor value is:

$$C = \frac{V_M I_M}{2\ \Delta V_{out}\ \omega\ V_{out}}$$

3) Values of the PFC components and parameters of the controllers

Inpout Voltage Vin (V)	230
Output Voltage Vout (V)	450
Frequency (Hz)	50
Pulsation (rad/s)	314,159265
Switching Frequency (Hz)	20000
Output Power (W)	3700

ΔV (%)	3
ΔV (V)	13,5

ΔI (%)	20
ΔI (A)	4,55007842

Current value	
Imax=2*Ps/Vin max	
Imax (A)	22,7503921

Current sensor gain Kcs	1
Max voltage carrier wave Vcmax (V)	10
Cutoff Fequency current loop (Hz)	2000
Cutoff Pulsation ωcl (rad/s)	12566,3706
Phase Margin M(F) (°)	45
Phase Margin M(F) (rad)	0,78539816

Voltage sensor gain Kvs	0,025
Cutoff Fequency voltage loop (Hz)	20
Cutoff Pulsation ωvl (rad/s)	125,663706
Time constant closed loop (1/ωvl) τcl (s)	0,00795775

Resistor Load Value	
R=Vout²/Pout	
R (Ω)	54,72973
Inductor Value	
L=Vout/(4*fswi*ΔI)	
L (mH)	1,2362424
Capacitor Value	
C=(Vin max*Iin max)/2*Δvout*ω*Vout	
C (mF)	1,9386775

Current PI controller	Kp + Ki/p
Kp=(Vcmax *L*ωcl)/(Kcs*Vout)	
Kp	0,345224
τ=(tan(M(Φ)))/ωcl	
τ	7,958E-05
Ki= Kp/τ	
Ki	4338,2129
Voltage PI controller	Kv+Ki/p
Kv(1+tv.p)/(tv.p)	
τ = (RC/2)	
τv	0,0530516
Kv=(4*Vout*τv)/(Kvs*R*Vin max*τcl)	
Kv	26,963428
Ki= Kv/τv	
Ki	508,24864

4) Efficiency of the OBC & Matlab script

The first thing to do was to determine the data of the components affecting the losses. For example, the diodes of the input rectifier bridge have conduction losses. These losses are function of the input current and of the drop voltage during the “on state”.

$$P_{cond} = V_F I_F$$

The problem is that the drop voltage is not constant and is depending, mainly, on the current. The first step is to determine the equation of the drop voltage according to the current in the diode. This is realized thanks to the datasheet component and Excel by the following method.

- Find the expression of the drop voltage function of the current in the datasheet.

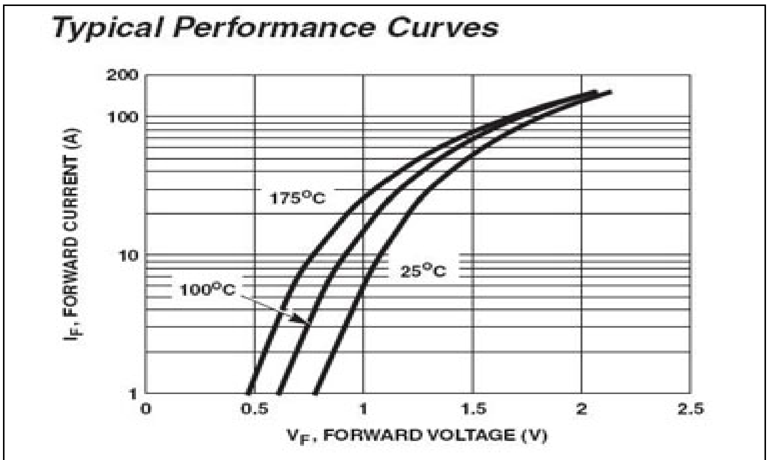
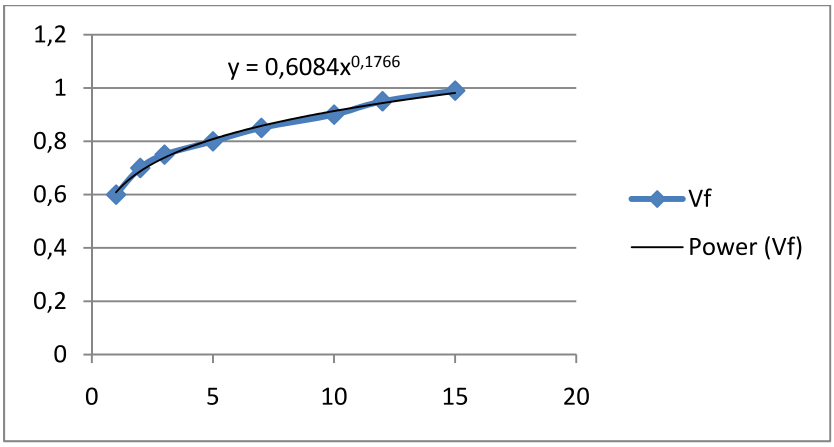


FIGURE 1. FORWARD CURRENT vs FORWARD VOLTAGE

- Fill in an excel table with points of the curve.

I_f (A)	1	2	3	5	7	10	12	15
V_f (V)	0,6	0,7	0,75	0,8	0,85	0,9	0,95	0,99

- Draw the curve and add a trend curve with Excel.



The expression of the trend curve gives the drop voltage function of the voltage.

Matlab script:

```
%On board charger losses

%Data
Ubat=400;                                               %Battery voltage
fswi=20000;                                             %Switching frequency
D=Ubat/450;                                             %Duty cycle
Vrrm=450;                                               %Maximum reverse voltage
diode

%loop
i=0;
I=0;
ii=0;
for k=1:0.1:16
   ii=ii+1;
   I(ii)=k;

%PFC
   %input diode bridge losses
     Vfib(ii)=0.608*((I(ii))^0.176);       %diode drop voltage
     Pcibd(ii)=2*Vfib(ii)*I(ii);           %diode conduction losses
     Psibd=0;                              %diode switching losses =0 a 50Hz
   %boost transistor
     Vcepfc(ii)=0.489*((I(ii))^0.356);     %transistor drop voltage
     Etot=7.8e-05*(I(ii))+0.05e-03 ;       %energy losses during turn phases
     Pcpfct(ii)=Vcepfc(ii)*I(ii);          %transistor conduction losses
     Pspfct(ii)=fswi*(Etot);               %transistor switching losses
     Pbt=Pcpfct+Pspfct;                    %transistor losses
   %boost diode
     Vfpfc(ii)=0.173*log(I(ii))+0.724;     %diode drop voltage
     Pcpfcd(ii)=Vfpfc(ii)*(I(ii));         %diode conduction losses
     Pspfcd(ii)=fswi*(-5e-11*(I(ii))
           +8e-9*(I(ii))+1e-7)*Vrrm/2;     %diode switching losses
     Pbd=Pcpfcd+Pspfcd;                    %diode losses
   %capacitor
     Pcappfc(ii)=0.059*(0.1*I(ii))^2;      %capacitor losses
   %inductor
     Pwindpfc(ii)=0.048*(I(ii))^2;         %inductor wire losses
     Pcindpfc(ii)=0.036*((I(ii))^2.031);   %inductor core losses
     Pindpfc=Pwindpfc+Pcindpfc;            %inductor losses
   %PFC losses
     Ppfc=Pcibd+Pbt+Pbd+Pcappfc+Pindpfc;   %losses in the PFC
end



j=0;
J=0;
jj=0;
for kk=1:0.1:16
   jj=jj+1;
   J(jj)=kk;
%____________________Duty cycle control_______________________%
%Full bridge
   %Transistors and freewheeling diodes
    Vfib(jj)=0.489*((J(jj))^0.356);        %transistor drop voltage
    Pcfbtdc(jj)=2*D*Vfib(jj)*(J(jj));      %transistor conduction losses
    Psfbtdc=0;                             %transistor switching losses (=0)
```

```
%Diode rectifier bridge
    Vffb(jj)=0.173*log(J(jj))+0.724;      %diode drop voltage
    Pcfbddc(jj)=(2*D*Vffb(jj)*(J(jj)))
         +4*(1-D)*((Vffb(jj)/2)*(J(jj)));%diode conduction losses
    Psfbd1dc(jj)=(-5e-11*(J(jj))
         +8e-9*(J(jj))+1e-7);             %Qrr function of the current
    Psfbddc(jj)=fswi*(2*Psfbd1dc(jj)*Vrrm/2)
         +fswi*2*(Psfbd1dc(jj)*Vrrm/2);   %diode switching losses
    Pdrdc=Pcfbddc+Psfbddc;                %diode rectifier bridge losses
   %Transformer
    Pwtransfdc(jj)=2*D*0.0163*(J(jj))^2;  %transformer wire losses
    Pctransfdc(jj)=0.705*(J(jj));         %transformer core losses
    Ptransdc=Pwtransfdc+Pctransfdc;       %transformer losses
   %Capacitor
    Pcapfilt(jj)=0.007*(0.1*J(jj))^2;     %capacitor losses
   %Inductor
    Pwindfilt(jj)=0.094*(J(jj))^2;        %inductor wire losses
    Pcindfilt(jj)=0.292*((J(jj))^1.924);  %inductor core losses
    Pindfilt=Pwindfilt+Pcindfilt;         %inductor losses
   %Full Bridge losses
    Pfbdc=Pcfbtdc+Pcfbddc+Psfbddc
         +Ptransdc+Pcapfilt+Pindfilt;     %losses in the full bridge
   %OBC losses
    Pobcdc=Ppfc+Pfbdc;                    %losses in the on-board charger

%____________________Phase shift control______________________%
%Full bridge
   %Transistor and freewheeling diode
    Vfib(jj)=0.489*((J(jj))^0.356);       %transistor drop voltage
    Pcfbtps(jj)=2*D*Vfib(jj)*(J(jj))
        +2*(1-D)*(Vfib(jj)*(J(jj)));      %transistor conduction losses
    Psfbtps=0;                            %transistor switching losses
   %Diode rectifier bridge
    Vffb(jj)=0.173*log(J(jj))+0.724;      %diode drop voltage
    Pcfbdps(jj)=(2*Vffb(jj)*(J(jj)));     %diode conduction losses
    Psfbd1ps(jj)=(-5e-11*(J(jj))
                +8e-9*(J(jj))+1e-7);      %Qrr function of the current
    Psfbdps(jj)=fswi*(2*Psfbd1ps(jj)*Vrrm/2);   %diode switching losses
    Pdrps=Pcfbdps+Psfbdps;                %diode rectifier bridge losses
   %Transformer
    Pwtransfps(jj)=2*0.0163*(J(jj))^2;    %transformer wire losses
    Pctransfps(jj)=0.705*(J(jj));         %transformer core losses
    Ptransps=Pwtransfps+Pctransfps;       %transformer losses
   %Capacitor
    Pcapfilt(jj)=0.007*(0.1*J(jj))^2;     %capacitor losses
   %Inductor
    Pwindfilt(jj)=0.094*(J(jj))^2;        %inductor wire losses
    Pcindfilt(jj)=0.292*((J(jj))^1.924);  %inductor core losses
    Pindfilt=Pwindfilt+Pcindfilt;         %inductor losses
   %Full Bridge losses
    Pfbps=Pcfbtps+Pcfbdps+Psfbdps
         +Ptransps+Pcapfilt+Pindfilt;     %losses in the full bridge
   %OBC losses
    Pobcps=Ppfc+Pfbps;                    %losses in the on-board charger
end
```

```
%Efficiency Duty Cycle Control
    %losses @16A
P16dc=Ppfc(151)+Pfbdc(floor(((3680/Ubat)*10)-9));
    %efficiency @16A
n16dc=((3680-P16dc)/3680*100)
…………………………………………………………………………………………………………………

…………………………………………………………………………………………………………………

%losses @6A
P6dc=Ppfc(51)+Pfbdc(floor(((1380/Ubat)*10)-9));
    %efficiency @6A
n6dc=((1380-P6dc)/1380*100);


%Efficiency Phase Shift Control
    %losses @16A
P16ps=Ppfc(151)+Pfbps(floor(((3680/Ubat)*10)-9));
    %efficiency @16A
n16ps=((3680-P16ps)/3680*100)
…………………………………………………………………………………………………………………

…………………………………………………………………………………………………………………

    %losses @6A
P6ps=Ppfc(51)+Pfbps(floor(((1380/Ubat)*10)-9));
    %efficiency @6A
n6ps=((1380-P6ps)/1380*100);

%Graph
hold on

figure(1)
%losses in each semiconductors
xlabel('Current (A)')
ylabel('Losses (W)')
plot(I,Pcibd,' r',I,Pbt,' b',I,Pbd,' m',I,Pcfbtdc,' k',I,Pdrdc,'
g',I,Pcfbtps,' y',I,Pdrps,' c')
title('Losses in the semiconductors function of the current')
text(1.5,28,'red      =  input rectifier bridge')
text(1.5,31,'blue     =  boost transistor')
text(1.5,22,'pink     =  boost diode')
text(1.5,37,'yello    =  inverter PS')
text(1.5,25,'black    =  inverter DC')
text(1.5,34,'cyan   =  output rectifier PS')
text(1.5,40,'green  =  output rectifier DC')

figure(2)
%Efficiency of the on board charger
plot(6,n6dc,' b+',7,n7dc,' b+',8,n8dc,' b+',9,n9dc,' b+',10,n10dc,'
b+',11,n11dc,' b+',12,n12dc,' b+',13,n13dc,' b+',14,n14dc,' b+',15,n15dc,'
b+',16,n16dc,' b+',6,n6ps,' r+',7,n7ps,' r+',8,n8ps,' r+',9,n9ps,'
r+',10,n10ps,' r+',11,n11ps,' r+',12,n12ps,' r+',13,n13ps,' r+',14,n14ps,'
r+',15,n15ps,' r+',16,n16ps,' r+')
title('Efficiency in the OBC function of the current')
xlabel('Current (A)')
ylabel('Efficiency (%)')
text(12,95.65,'red = phase shift control')
text(12,95.8,'blue = duty cycle control')
```