Theoretical Design of a Single Phase 2kW Power Supply with 1.5kV DC Output

Per Mossadeghi Björklund
Abstract

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This 15hp degree project in electrical engineering is carried out at ScandiNova Systems AB in Uppsala, Sweden. The company constructs pulse generators used in areas such as sterilization or radiation therapy. As a step in development they want to design their own power supply for one of their generators instead of purchasing it from an external supplier.

This paper is about how to make a theoretical design of a power supply that can use worldwide single-phase AC line input and deliver ~1500V DC output.

The paper has a power loss and harmonic theory part and a method part for power inductor calculation etc.

The result is a power supply seen from the AC line starting with a second order EMI-filter, a full-wave rectifier bridge, a PFC (Power Factor Correction) boost converter based on the controller UCC28180 from Texas Instruments and a SLR (Series Loaded Resonant) half-bridge converter with control and feedback produced by the company itself. Associated energy efficiency calculations, drawing, component lists and physical design compared to the size of the current power supply and components made in SolidWorks is the design output.

The conclusion is that the goal parameters are theoretically fulfilled and realistic but that experimental testing is needed for optimizing the design and control. That is why some of the parameters are changeable for best modifiability later in the product design.
Sammanfattning


Denna uppsats handlar om hur man gör en teoretisk konstruktion av ett nätaggregat som kan användas i hela världen med enfas växelström på linjeingången och leverera ~1500V på DC-utgången.

Uppsatsen har en teoridel om effektförluster och harmoniska övertoner och en metod del om beräkning av induktor etc.

Resultatet är ett nätaggregat sedd från AC linjeingång som börjar med ett andra ordningens EMI-filter, en helvägslikerktare, en PFC (Power Factor Correction) boost omvandlare baserad på styrenheten UCC28180 från Texas Instruments och en SLR (Series Loaded Resonant) halvbrygga med styrning och återkoppling som produceras av företaget självt. Energieffektivitetsberäkningar, ritningar, komponentlistor och fysisk utformning jämfört med storleken på det aktuella nätaggregatet och komponenter tillverkade i SolidWorks är projektets utkomst.

Slutsatsen är att målparametrarna teoretiskt är uppfyllda och realistiska, men att experimentell testning behövs för att optimera design och kontroll. Det är därför en del av parametrarna är utbytbara för bästa modifierbarhet senare i produktdesignen.
Acknowledgement

This thesis in electrical engineering is dedicated to my beloved wife and son that made this education possible for me to manage because of their energy and patience.

I would like to thank the people who support me during this work:
Ulf Holböll, Klas Elmquist and Magnus Graas at ScandiNova Systems AB.
Markus Gabrysch, Nóra Masszi and Kjell Pernestål at Uppsala University.

I would also like to thank my parents for all their help.

Per Mossadeghi Björklund
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Dec -14
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1 Introduction

1.1 Background
This 15hp bachelor work in electrical engineering is carried out at ScandiNova Systems AB in Uppsala, Sweden. The company was founded in 2001 and their business is to build high-quality pulse generators, so called modulators. These are used in several areas such as radiation therapy, industrial radiography equipment, sterilization, and RADAR scientific accelerator systems. The specifications of the pulses range between 1 kV – 500 kV.

1.2 Aim
Today the company produces its own power supplies to their larger modulators, but purchases commercial power supplies for the smaller single-phase pulse generators called M0.5. Their aim is to manufacture these by themselves in the company and this work is a theoretical study to find some alternative solution for them for the future. High efficiency is a key parameter since less power not only saves money directly but also indirectly by decreasing cooling costs for the customer.

The company also needs to implement active PFC (power factor correction) in this device to fulfill the EN61000–3–2 power factor regulation and reduce the line distortions and losses that a low power factor causes by moving a lot of reactive energy between the source and the sink.

1.3 Goal
The goal parameters for the power supply are:

- EN55011 standard in conducted disturbances in the 150 kHz – 30 MHz range
- EN61000–3–2 standard power factor
- 95% energy efficiency
- Single phase
- 1000W output with 100V AC\(_{\text{RMS}}\) input (Japan etc.)
- 1500W output with 120V AC\(_{\text{RMS}}\) input (North America etc.)
- 2000W output with 230V AC\(_{\text{RMS}}\) input (Europe etc.)
- Worldwide standard line input
- \(\sim1500\)V DC output

1.4 Boundaries
This work only contains the power supply and no other part of the modulator. The project is only a theoretical work without the aim to build and test the designed converter. A design for decreasing the radiated distortions that might appear is excluded as well as PCB layout.
1.5 Design

The power supply’s design (figure 1) is chosen according to energy efficiency, component price and size of the current power supply.

The power supply is made of five parts which is an EMI-filter (section 2.4), two full-wave rectifiers (section 2.5), a PFC boost converter with associated PFC controller (section 2.6) and a SLR half-bridge converter (section 2.7). The voltage after the boost converter is ~390V and after the half-bridge ~1500V. The boost converter has a switching frequency of 65 kHz and the half-bridge 60 kHz to avoid interaction.

The step-up from line voltage was necessary and a boost converter with built-in PFC seemed to be the best choice when it solves the world wide line input, PFC and make the step up to 390V in one solution. The company already had a fine working half-bridge solution. The benefits with active PFC according to line losses versus energy efficiency of the power supply can be discussed but should not affect when it is a way of handling the switch a little different from a pure boost converter.
Figure 1: Proposed power supply showing clearly the five converter stages with the AC input in the left and the DC output on the right.
2 Theory

2.1 Origin of dominant power losses in a power supply

2.1.1 Wire

The power loss in a conductor is defined as:

\[ P = I^2 \cdot R \text{ where } R = \rho \cdot \frac{l}{A} \]  \hspace{1cm} (1), (2)

From the definition (1) the power loss in a wire depends on how much current passes \((I)\) and the resistance \((R)\) which depends on the length \((l)\), the cross sectional area \((A)\) and resistivity \((\rho)\).

2.1.2 Magnetic core

The magnetic core has two main losses. One is hysteresis which occurs when the magnetic field in the core is changing and the electrons have to move and align according to the new and always changing magnetic field. The material’s BH-curve tells about the losses of one cycle that are proportional to the area of the closed loop. The area is material dependent and the losses will increase with higher frequencies when moving through the curve at a higher rate per time unit. The other type of loss is eddy current losses which occur when a straight magnetic field is varying in time and forces the electrons to circulate around it. The currents produce heat through resistance in the material like the wire losses. A way to decrease the circulating currents is to laminate the core in thin slices with insulating coating.

2.1.3 MOSFET

In the MOSFET there are two type of losses. One is conduction loss like wire loss and occurs because of the current and the resistivity from drain to source which is named \(R_{DS(ON)}\) in the datasheets. The second one is switching loss which can be defined as:

\[ P_{sw} = f_{sw} \cdot (0.5 \cdot V_{out} \cdot I_{max}(t_{rise} + t_{fall}) + 0.5 \cdot C_{oss} \cdot V_{out}^2) \]  \hspace{1cm} (3)

Where \(t_{rise}\) and \(t_{fall}\) is the rise and fall time and \(C_{oss} = C_{gd} + C_{ds}\). \(C_{oss}\) is referred to as the transistor’s small signal output capacitance with the gate and source terminals shorted. Figure 2 describes the power loss of a MOSFET during one switching cycle.
2.1.4 Diode
The diode losses are based on the forward voltage drop ($V_F$) and the reverse recovery charge ($Q_{RR}$) according to (4).

$$P = V_F \cdot I_{\text{max}} + 0.5 \cdot f_{\text{sw}} \cdot V_{\text{out}} \cdot Q_{RR}$$  \hspace{1cm} (4)

If a voltage is applied across a silicon pn-diode there will be a voltage drop over the pn-junction of around 0.7V. The conduction losses are calculated as $V_F$ multiplied with the current ($I_{\text{max}}$). The reverse recovery charge ($Q_{RR}$) determines the time it takes for the diode to switch off.

2.2 Origin of dominating harmonic disturbances from a power supply
Disturbances in electrical systems often occur in circuits where other waveforms then the regular line-frequency sine wave are present. According to Fourier’s theorem all periodic waveforms $f(t)$ with period $T$ can be built of sinusoidal waves of integer multiple of the fundamental frequency (5).

$$f(t) = \frac{a_0}{2} + \sum_{n=1}^{\infty} \left( a_n \cos \frac{n2\pi t}{T} + b_n \sin \frac{n2\pi t}{T} \right)$$  \hspace{1cm} (5)

In a switch-mode power supply the square wave is a common shape because of the switching. The Fourier series of an odd square wave with unity amplitude looks like (6).

$$f(t) = \frac{4}{\pi} \sum_{n=1,3,5,\ldots}^{\infty} \left( \frac{1}{n} \sin \frac{n\pi t}{T} \right)$$  \hspace{1cm} (6)

It shows the construction of the square wave as an infinite sum of sine waves built on each other but with decreasing amplitude.
2.3 Definition of power factor

The input power factor is generally defined like (7).

\[
Power\ factor\ (PF) = \frac{P (average/real\ input\ power)}{|S| (apparent\ input\ power)} = \frac{P_{dc}}{V_{RMS}I_{RMS}} \tag{7}
\]

Where \( P_{dc} \) is defined like (8).

\[
P_{dc} = V_{1,RMS} \cdot I_{1,RMS} \cdot \cos(\varphi) \tag{8}
\]

Where \( V_{1,RMS} \) and \( I_{1,RMS} \) is the fundamental 50Hz or 60Hz sinusoidal waveform and \( \cos(\varphi) \) the displacement angle between the two. Because \( V_{RMS} \) and \( V_{1,RMS} \) are the same they can be removed and the remaining is the power factor equation when current harmonics occurs in a system (9).

\[
PF = \frac{I_{1,RMS}}{I_{RMS}} \cdot \cos(\varphi) \tag{9}
\]

The ratio between the RMS of the fundamental current and the total current is called distortion power factor (10).

\[
Distortion\ power\ factor = \frac{I_{1,RMS}}{I_{RMS}} \tag{10}
\]

The power factor (9) was (for the case of a sinusoidal voltage) the product of the “distortion power factor” and the “displacement power factor” \( \cos(\varphi) \) where 0 is the case with no real power transferred and 1 is the case with only real power transferred. Also a PF of less than 0 is possible when energy flows (in average) from the load to the source, i.e. during regenerative braking of a motor.

![Figure 3: PFC demonstration which shows the line voltage and the absolute value of the input current with and without PFC](image)

The power factor correction tries to make a non-sinusoidal input current look as similar as the sinusoidal input voltage as possible like the red curve (I_in with PFC) in figure 3 demonstrates to obtain a high power factor. In high power applications a high power factor is necessary when the electric grid operators want the costumers to fulfill regulations such as the EN61000-3-2 to maintain a high quality line with low reactive losses and disturbances. Even for the costumer a high power factor is beneficial since smaller wires and lower level fuses can be used.
2.4 EMI-filter

EMI (Electromagnetic interference) is disturbance in one or two different devices simultaneously caused by conducted and/or radiated emissions from each other. The EMI range is set to 10 kHz – 30 MHz by conduction through wires and between 30 MHz – 1 GHz by radiation.

There are two different kinds of noise. One is common mode (CM) disturbance which is measured on both the line and the neutral with ground as reference. The main reasons for it are high rates of \( \frac{dv}{dt} \) when switching and parasitic capacitances to ground. The other is differential mode (DM) disturbance which is measured on the line with the neutral as a reference. The reason is the high rate of \( \frac{di}{dt} \) because of spikes in a switching MOSFET or diode.

![Figure 4: EMI filter example](image)

To remove unwanted conducted disturbances an EMI-filter can be used. The radiated ones are removed by choosing appropriate design and shielding of the device but this is not discussed in this paper.

In figure 4, \( L_{\text{CM}} \) is a common-mode choke and \( L_{\text{DM}} \) is a differential-mode choke. \( C_{x1} \) and \( C_{x2} \) are differential-mode capacitors and the two \( C_y \) are a common-mode capacitors. \( L_{\text{DM}} \) has two opposite windings and \( L_{\text{CM}} \) has two identical windings round the same core.

There are a several standards how good a device should be on suppressing the noise. In this work the EN55011 standard (figure 5) was chosen as reference because the power supply fits the description:

- “Industrial, scientific and medical (ISM) equipment.” [15]

Classification:

- “Class B covers devices for usage in domestic establishments that are directly connected to a low voltage power supply network, which supplies domestic environment.” [15]

Requirements:

- “Conducted emission requirements for mains ports and telecommunication ports must be met in frequency range 150 kHz – 30 MHz.” [15]
2.5 Full-wave rectifier

A full-wave rectifier contains four diodes with the needed voltage rating.

When an AC voltage source is connected to the full wave rectifier (figure 6) a diode pair only conducts for positive voltage and blocks for negative voltage while the other diode pair does the opposite. In order not to reduce the PF, a moderate sized smoothing capacitor is used that allows us to receive a voltage ripple of around 6% [14] to have a reference when shaping the input current for high power factor.
2.6 Boost converter with active PFC

A boost converter is a step-up DC-to-DC converter and as the name reveals the output voltage is higher than the input voltage. Figure 7 shows the different modes the boost converter can be in and its four main components, a boost inductor (L1), a transistor, a diode (D1) and a storage capacitor (C1).

Figure 7: Boost converter in operation [2]

In the first quadrant of figure 7 the basic boost converter circuit is shown. In the second quadrant the gate voltage turns high and a current is now charging the boost inductor with energy. When the gate voltage turns low again in the third square the built up energy rushes out from the boost inductor towards the storage capacitor and the load. How much energy is delivered depends on the size of the inductor and the PWM timing at the gate. In the fourth and last stage the boost inductor is reloaded and the circuit is divided into two sub-circuits by the blocking diode. The load gets in this stage all its energy from the storage capacitor until a new cycle from quadrant three begins.

The two most difficult components to decide about are the MOSFET and the boost inductor. It is very important that the MOSFET has as low $R_{DS(ON)}$ as possible because a high drain current $I_{RMS}$ would otherwise result in huge power losses (dissipated as heat in the device). When designing a boost inductor properties like core material, cable size and winding have to be considered.

The power factor correction is achieved by the UCC28180 controller chip from Texas Instruments. The chip has an 8-pin connector and fulfills the goal parameters of the system. The controller has an integrated gate drive and the switching frequency can be set between 18 kHz – 250 kHz depending on an external resistor. The switching frequency is of high
importance when it comes to switch losses. Higher frequency gives higher losses but stands against more bulky components and in the end the size of the power supply.

Power factor correction is a way of suppressing the harmonics close to the fundamental (line) frequency if the application has an uneven current consumption. The EMI-filter aims at removing the disturbances at high frequencies and the PFC-circuit tries to remove low frequency input current harmonics in order to decrease power line losses.

The selected UCC28180 controller chip operates under Continuous Conduction Mode (CCM). It means that the inductor current flow never stops but stays on a high level almost in every condition. This mode is popular in higher power levels as it has low peak and RMS current. This reduces stress in the MOSFET, diode and inductor. Designing the EMI-filter and boost inductor is easier because of the continuous current and the fixed switching frequency [6].

The control- and feedback system is of a newer predictive average current mode type which decreases the external number of components and complexity. The PWM modulator gets its feedback from the voltage drop over the sense resistor and the 5V comparative voltage from the output voltage. The PWM stage generates a periodic ramp function (figure 8) and makes the gate voltage high whenever the ramp voltage exceeds the $V_{ICOMP}$ voltage and low for the opposite. The ramp/$V_{ICOMP}$ intersection determines the off timing, and hence duty cycle off ratio. Since $DutyOFF = \frac{V_{IN}}{V_{OUT}}$ by the boost topology equation, $V_{IN}$ is sinusoidal and $V_{ICOMP}$ is proportional to the boost inductor current. It follows that the control loop forces the average inductor current to follow the input voltage wave-shape to maintain boost regulation [14]. In this way the controller can match the input sine wave without separate line sensing input.

![PWM generation](image)

**Figure 8: PWM generation [14]**

### 2.7 Series-loaded resonant half bridge converter

The step-up after the boost converter is a SLR (Series-Loaded Resonant) half-bridge converter (figure 9) with a high frequency transformer followed by a (second) full-wave rectifier. The name series-loaded is because the load always appears to be in series with the LC resonant tank. The maximum output voltage is the transformer ratio times half of the
rectified line voltage. The transformer’s primary side leakage inductance is utilized as the resonating inductance and in that way no additional inductance is needed.

![Diagram: Series-loaded resonant half-bridge converter](image)

Figure 9: Series-loaded resonant half-bridge converter [5]

This converter operates in discontinues mode which means that the current in the resonating capacitor drops to zero every half cycle of the operation (figure 10).

![Diagram: Current in, and voltage across C_res](image)

Figure 10: Current in, and voltage across C_res [5]

During one conduction period of one MOSFET the voltage swings from \(-V/2\) to \(+V/2\) (in our case from -195V to +195V). During that time the capacitor transfers one packet of charge, \(\Delta Q = CV\) and gives us an average current of \(I_{avg} = 2VCf_{sw}\). In this mode the MOSFETs are switched on during zero current conditions and off during zero current and zero voltage conditions which mean reduced stress on the switches and also lower switching losses but they might have higher conduction losses through higher currents. The control and feedback system of the switches is in this case a product of ScandiNova Systems AB and confidential.
3 Method and design
This is the method and design section which is showing the approaches of how to calculate this power supply’s components and energy losses.

3.1 How to calculate an EMI-filter
A device called LISN (Line Impedance Stabilization Network) is connected between the AC input and the equipment under test. The main reasons to use it is that it filters disturbance of higher frequency than line frequency, creates a known impedance for the equipment under test and gives a connection where you can connect a spectrum analyzer (figure 11).

Figure 11: Schematic of EMI-filter analysis [10]

The common mode noise will see two 50Ω resistors in parallel while differential noise sees two 50Ω resistors in series. The spectrum analyzer receives the total conducted noise from LISN and it needs to be separated into CM and DM components.

- \( V_{\text{LINE}} = V_{\text{CM}} + V_{\text{DM}} \)
- \( V_{\text{NEUTRAL}} = V_{\text{CM}} - V_{\text{DM}} \)
- \( V_{\text{CM}} = (V_{\text{LINE}} + V_{\text{NEUTRAL}})/2 \)
- \( V_{\text{DM}} = (V_{\text{LINE}} - V_{\text{NEUTRAL}})/2 \)

There is a good procedure how to accomplish a filter that gives the wanted result [10].

- Measure the differential- and common mode noise levels
- Determine the differential- and common mode attenuation requirements using
  - \( (V_{\text{req, CM}})dB = (V_{\text{CM, measured}})dB - (V_{\text{limit}})dB + 6\ dB \) safety margin
  - \( (V_{\text{req, DM}})dB = (V_{\text{DM, measured}})dB - (V_{\text{limit}})dB + 6\ dB \) safety margin
- Determine the corner frequency by drawing a 40 dB/Dec slope which is tangent to the required attenuation for DM and CM at 150kHz (figure 12)
- Calculate component values
Because of the high complexity for any SPICE program to calculate the nodal voltage and currents with the extremely small time step needed to calculate the noise in the frequency domain beginning at 150 kHz an estimation was based on [11]. It says that the general corner frequency for CM noise is 28 kHz and for DM noise 20.5 kHz.

CM components:

C_y is limited to 5400pF [10] because of safety rules regarding time of risk for electric shock from stored charge if power supply is disconnected from AC line. Choose C_y to the largest value 5.4nF for the smallest choke according to (11).

\[
L_c = \left(\frac{1}{2\pi f_{R, CM}}\right)^2 \cdot \left(\frac{1}{2C_y}\right) = \left(\frac{1}{2\pi \cdot 28000}\right)^2 \cdot \left(\frac{1}{2 \cdot (5.4 \cdot 10^{-9})}\right) = 3\text{mH}
\]  

(11)

According to (11) L_c will be 3mH. To reduce cost and size of the filter manufacturers often use the common mode inductor’s leakage inductance L_{leak} as differential mode inductor. Without measurements the leakage inductance is chosen to 1.25% of 3mH, so L_{leak} then is 41.25µF. Normally L_{leak} is around 0.5 – 2 % of the primary inductance [10].

DM components:

\[
C_{x1} = C_{x2} = C_{DM} = \left(\frac{1}{2\pi f_{R, DM}}\right)^2 \cdot \left(\frac{1}{L_{leak}}\right) = \left(\frac{1}{2\pi \cdot 20500}\right)^2 \cdot \left(\frac{1}{41.25 \cdot 10^{-6}}\right) = 1.46\mu F
\]  

(12)

According to (12) C_{DM} is 1.46µF and a filter that meets the expectations can be built.
3.2 Boost converter with active PFC

The calculation of the boost converter components are made with reference to the TI UCC28180 datasheet [14] and the UCC28180 Microsoft Excel design guide. Table 1 is showing the basic parameters the calculations of the boost converter are based on. The input voltage is varying within the values set by the expected environment of the power supply.

Table 1: Table 1 is showing the basic parameters the calculations of the boost converter are based on.

<table>
<thead>
<tr>
<th></th>
<th>Japan</th>
<th>USA</th>
<th>Europe</th>
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</thead>
<tbody>
<tr>
<td></td>
<td>min</td>
<td>std</td>
<td>max</td>
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<td>Input voltage [V_rms]</td>
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<td>100</td>
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<td>Max output power [W]</td>
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<tr>
<td>Output voltage [V]</td>
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<tr>
<td>Max output current [A]</td>
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<td>Expected power factor</td>
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<td>0.99</td>
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<tr>
<td>Expected efficiency</td>
<td>0.95</td>
<td>0.95</td>
<td>0.95</td>
</tr>
</tbody>
</table>

3.2.1 Design of boost inductor for CCM operation

The calculation of the boost inductor starts with the selection of basic properties:

The switching frequency is 65 kHz because of a good compromise between the size of the boost inductor and the switching losses.

The material of the core is made of MPP material because of its low losses and good inductance stability in DC bias conditions.

The inductance is chosen with the help of equation 28 in [14] with a current ripple of 3.33A which is 15-25% of max input current which was recommended [14]. The inductance is 450µH.

Maximum inductor current is 22.9A and occurs when having a low (108V) input voltage at 1500W max output according to equation (6) and (26) in [14].

With these parameters, we can calculate the LI²-product which is a measure of the energy storage and needs to be fulfilled for proper function of the inductor (13).

\[ LI^2 = 236mHA^2 \]  \hspace{1cm} (13)

According to figure 13 the smallest core from Magnetics that fits the LI² product can be selected and it seems that the core 55735 is a good choice.
The datasheet of the core gives the following information. Permeability is 26µ and $A_L = 88 \pm 8\%$ where the $A_L$ (nH/turns$^2$) value is a nominal value of inductance.

$$A_{L_{min}} = 88\text{nH/turns}^2 - 8\% = 81\text{nH/turns}^2$$  \hspace{1cm} (14)

Then the number of turns (N) that is needed for the required inductance is calculated (15).

$$N = \frac{L_{req}}{\sqrt{A_{L_{min}}}} = 74.5 \text{ turns}$$  \hspace{1cm} (15)

Now the DC bias level of the core can be calculated. $I_e$ is the average core path in centimeters.

$$H = \frac{N \cdot I}{I_e} = \frac{74.5 \text{ turns} \cdot 22.9\text{A}}{18.4 \text{ cm}} = 92.77 \text{ A-turns/cm}$$  \hspace{1cm} (16)
The next step is to correct the DC bias level according to the core environment (figure 14).

\[ N = \frac{74.5 \text{ turns}}{0.86} = 87 \text{ turns} \]  

This gives a new \( H = 108.3 \text{ A-turns/cm} \) and a correction factor of the \( A_L \)-value of 0.8.

\[ L = N^2 A_L = 490 \mu\text{H} \]  

According to the new \( A_L = 81 \cdot 0.8 = 64.8 \text{ nH/turns}^2 \) the environmental adjustments are ready and give an inductance of 490\( \mu\text{H} \) versus the required 450\( \mu\text{H} \) which means OK.

The “right” wire size can now be chosen by a wire table [8]. A thicker wire has lower resistance but results in a larger and heavier inductor. An AWG 10 magnet wire was chosen that has a resistance of 3.28m\( \Omega \)/m.

The fill factor is a measure of how much the center of the toroid is taken by winding. The recommendation is 30 – 45 % [8].

\[ \text{fillfactor} = \frac{N A_w}{W_a} = \frac{87 \text{ turns} \cdot 0.056 \text{ cm}^2}{15 \text{ cm}^2} = 31.5\% \]  

Where \( A_w \) is the cross sectional area of the wire and \( W_a \) is the core window area. The recommended 30 – 45 % fill factor is fulfilled with a good capability wire. As a test the magnetic flux density was calculated to know if the saturation limit was close (20).

\[ B = \frac{N I_0 \mu_r}{l} \]  

The result was 0.35 T which fulfills the 0.75 T saturation limit for the MMP material.
Loss calculation:

According to figure 15 the magnetizing force $H_{AC\ min} = 93 \ A\text{-turns/cm}$ gives a flux density equal to 0.27T and $H_{AC\ max} = 108 \ A\text{-turns/cm}$ gives a flux density of 0.33T.

\[
B_{pk} = \frac{\Delta B}{2} = 0.03T
\]  

(21)

(22) is the core loss density formula where a, b and c are constants from curve fitting.

\[
P_L = a \cdot B_{pk}^b \cdot f^c = 70.83 \cdot 0.03^{2.34} \cdot 65^{1.65} = 19\text{mW/cm}^3
\]  

(22)

Multiplied with the core volume (23).

\[
P_{fe} = P_L \cdot l_e \cdot A_e = \frac{0.019\text{W}}{\text{cm}^3} \cdot 18.4\text{cm} \cdot 4.97\text{cm}^2 = 1.74\text{W}
\]  

(23)

The core losses are 1.74W.

Wire resistance calculation:

\[
R_{DC} = (MTL) \cdot N \cdot \frac{R}{l} = 11.9\text{cm/turn} \cdot 87\text{turns} \cdot \frac{0.0000328\Omega}{\text{cm}} = 0.034\Omega
\]  

(24)

Where MTL stand for “mean turn length” which is a winding factor depending on core window fill factor. Multiplied with $I_{RMS}^2 = 15.01^2\text{A}$ this give us maximum 7.66W in wire losses.

### 3.3 Series-loaded resonant half bridge converter

Concerning the half-bridge, the first thing to decide about is which transformer ratio is needed to reach 1500V. From 195V that will be a minimum ratio of 7.7:1. 7.7 is used because of the higher primary winding currents that would occur if a higher difference in ratio was chosen.
The switching frequency is set to be 60 kHz and the resonance frequency is set to be 60 kHz/0.85 = 70.588 kHz. This is to operate the converter in discontinues current mode as wanted in order to guarantee soft-switching of the power transistors. A resonance frequency ~15% above switching frequency is recommended and used in other ScandiNova Systems converters [5]. 60 kHz was chosen to separate it from the boost converters 65 kHz as a quest to decrease distortion.

\[ C_{res} = \frac{N_{ratio}P_{out}}{2V_{DCin}V_{load}f_{switch}} = \frac{7.7\cdot2000W}{2\cdot390V\cdot1500V\cdot60000Hz} = 0.219 \mu F \] (25)

\[ L_{res} = \frac{1}{(2\pi f_{res})^2C_{res}} = \frac{1}{(2\pi\cdot70588Hz)^2\cdot0.219\mu F} = 23.17 \mu H \] (26)

\( V_{DCin} \) is the full DC voltage (390V) input. For this converter the resonance capacitor is 0.219 \( \mu F \) and the leakage inductance of the transformers primary winding has to be most 23.17 \( \mu H \) (25), (26).

\[ I_{avg\ pri.} = 2C_{res}V_{DCin}f_{switch} = 2 \cdot 0.219 \mu F \cdot 390V \cdot 60kHz = 10.26A \] (27)

\[ I_{peak} = \frac{\pi C_{res}V_{DCin}}{2T_1} = \frac{\pi \cdot 0.219 \mu F \cdot 390V}{2 \cdot \left(\frac{1}{60kHz}\right)} = 18.97A \] (28)

\[ I_{RMS} = \frac{I_{peak}}{\sqrt{2}} \sqrt{\frac{2T_1}{T_3}} = \frac{18.97A}{\sqrt{2}} \cdot \sqrt{\frac{2}{\left(\frac{1}{60kHz}\right)^2}} = 12.37A \] (29)

According to equation (27), (28), and (29) the primary currents are calculated and can then be divided by 7.7 to have the secondary currents. \( T_1 \) is the time of one delivered charge packet \( Q \) and \( T_3 \) is the period time of the switching frequency (figure 10).

### 3.3.1 How to measure primary leakage inductance

To find the transformers leakage inductance a so called short circuit test is made on the transformer.

Figure 16: Transformer short circuit test [13]

Figure 16 shows the connection diagram for the short circuit test. In this case the high voltage side of the transformer is short circuited and a wattmeter (W), voltmeter (V) and amperemeter (A) are connected on the low voltage side of the transformer. A voltage is applied to the low voltage side and increased from zero until the amperemeter reading is equal to the rated current of the transformer.

The amperemeter reading gives the primary equivalent of full load current (\( I_{sc} \)). The voltage applied for full load current is very small compared to the rated voltage. Hence, core loss due to small applied voltage can be neglected, (\( R_p \) and \( X_p \)). The wattmeter reading can be taken as...
copper loss in the transformer primary windings. Therefore, \( W = I_{sc}^2 R_{eq} \), where \( R_{eq} \) is the equivalent resistance in the windings. Because \( Z_{eq} = V_{sc} / I_{sc} \), the equivalent reactance of the transformer can be calculated from the formula \( Z_{eq}^2 = R_{eq}^2 + X_{eq}^2 \).

### 3.3.2 How to design a HF transformer

The calculation of the transformer starts with the \( W_Ac \) product where \( A_c \) is the cross sectional area of the core and \( W_a \) is the window area of the core. It tells about the cores power handling capacity. According to [7]

\[
W_Ac = \frac{P_{out} \cdot D_{cma}}{K_t \cdot B_{max} \cdot f_{res}} = 8.095 \text{cm}^4 \quad (30)
\]

Where \( D_{cma} \) is set to recommended 500 circular mils/amp [7], \( B_{max} \) to 1250 gauss (figure 17) and \( K_t \) is a constant for the half-bridge topology [7]. The \( W_Ac \) product of 9cm\(^4\) which is the closest larger value gives Magnetics core 0P45716U according to [7]. It is the smallest core according to the power handling capacity that can be chosen [7].

![FLUX DENSITY VS. FREQUENCY P MATERIAL](image)

**Figure 17: Frequency versus magnetic flux density [7]**

Figure 17 shows the maximum core flux density according to the frequency of operation for which the core losses become 100mW/cm\(^3\) and with a maximum temperature rise of 25°C. Then the necessary number of turns on the primary and secondary side were calculated (31), (32).

\[
N_{pri} = \frac{V_{pri}}{4 \cdot B_{max} \cdot A_c \cdot f_{res}} = \frac{195V}{4 \cdot 0.125T \cdot 0.000171 \text{m}^2 \cdot 70588 \text{Hz}} \approx 33 \text{ turns} \quad (31)
\]

\[
N_{sec} = \frac{U_{sec}}{U_{pri} N_{pri}} = \frac{1500V}{195V} \cdot 33 \text{ turns} \approx 254 \text{ turns} \quad (32)
\]
This gives a primary winding of 33 turns and a secondary winding of 254 turns. If a primary winding wire with a cross sectional area of 5.6mm² (American Wire Gauge, AWG 10) and a secondary winding wire with a cross sectional area of 0.754mm² (AWG 19) the core window fill factor can be obtained.

\[ KW_a \geq N_p A_{wp} + N_s A_{ws} \]  \hspace{1cm} (33)

Where \( K \) is a constant of maximum recommended fill factor 0.6 (60%) of the core window area \( W_a \).

\[ 0.6 \cdot 8.618 \text{cm}^2 \geq 33 \cdot 0.056 \text{cm}^2 + 254 \cdot 0.00754 \text{cm}^2 \]  \hspace{1cm} (34)

\[ 5.17 \text{cm}^2 \geq 3.76 \text{cm}^2 \]  \hspace{1cm} (35)

\[ 3.76 \text{cm}^2 / 8.618 \text{cm}^2 = 44\% \text{ fill factor} \]  \hspace{1cm} (36)

A fill factor of 44% is within the recommended 60% and OK. According to figure 17 this example is based on a core design with losses of 100mW/cm³ and according to the datasheet the core volume is 27.9 cm³ and this give a power loss of 2.79W.
4 Results and discussion

Most of the design parameters were made by hand according to [14], but with a little help of TI’s UCC28180 design calculator in Microsoft Excel. A lot of time the design calculator could not be used because of wrong calculated currents when it was chosen to have different maximum output power according to different line voltages.

This project was not reasonable to simulate because of huge time consuming analysis and complex control systems but the work can be considered realistic.

4.1 EMI-filter

All the components to build the EMI-filter are found in table 2.

<table>
<thead>
<tr>
<th>Part:</th>
<th>Name:</th>
<th>Package:</th>
<th>Value:</th>
<th>Cost:</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cx1</td>
<td>1.46µF</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>LCm</td>
<td>Wurth</td>
<td>3mH</td>
<td></td>
<td>125 sek</td>
</tr>
<tr>
<td>LDm</td>
<td>41.25µH</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Cx2</td>
<td>1.46µF</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>2 x Cy</td>
<td>5.4nF</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Sum:</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

4.2 Full-wave rectifier

The first full-wave rectifier turns a lot of energy into heat because of the low voltage and the high currents. The maximum average output current of the bridge is 11.54A in Japan, 13.52A in the USA and 9.36A in Europe. The rectifier diodes have a voltage drop of 1.1V at these currents. It gives a power loss of 25.4W, 29.75W and 20.6W. Even if the voltage drop is 1V it just makes a small difference. Table 3 shows the component list of the full-wave rectifier.

<table>
<thead>
<tr>
<th>Part:</th>
<th>Name:</th>
<th>Package:</th>
<th>Value:</th>
<th>Cost:</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rectifier bridge 1</td>
<td>IXYS VBO40-08NO6</td>
<td>SOT-227B (miniblock)</td>
<td>V_{RRM} = 800V</td>
<td>145 sek</td>
</tr>
<tr>
<td>Sum:</td>
<td></td>
<td></td>
<td></td>
<td>145 sek</td>
</tr>
</tbody>
</table>
4.3 Boost converter with active PFC

The boost converter was chosen because of the amount of examples of good feasibility with implementation of PFC. The microcontroller UCC28180 was chosen because of its simplicity and good reviews on its precursors (UCC28019 and UCC28019A). It is also meeting the demands of the converter and the switching frequency is changeable for best modifiability later in the work if the frequency interferes with the other switches when testing. A calculation on a core with 62mm (0055615A2) outer diameter that could work was made but came up with a fill factor of above 60%. The bigger core is better when it comes to modification opportunities when litz wire might be needed because of high skin effect in the magnet wire.

It is difficult to know if the power factor correction really helps to decrease the general energy consumption compared it was not used. But for sure the power supply was in need of a step-up from the line voltage anyway and PFC is only a way to handle the switch a little different. Probably the same losses should have occurred anyway.

Table 4 shows the component list of the PFC boost converter.
### Table 4: Component list for PFC boost converter

<table>
<thead>
<tr>
<th>Part</th>
<th>Name</th>
<th>Package</th>
<th>Value</th>
<th>Cost</th>
</tr>
</thead>
<tbody>
<tr>
<td>C&lt;sub&gt;IN&lt;/sub&gt;</td>
<td>Film Capacitor, X2</td>
<td></td>
<td>0.68µF (265V&lt;sub&gt;RMS&lt;/sub&gt;)</td>
<td></td>
</tr>
<tr>
<td>L&lt;sub&gt;BST&lt;/sub&gt;</td>
<td>Mag-Inc. 0055735A2</td>
<td>Toroid</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Q&lt;sub&gt;BST&lt;/sub&gt;</td>
<td>IXYS 75N60C</td>
<td>SOT-227B (miniblock)</td>
<td>V&lt;sub&gt;DSS&lt;/sub&gt; = 600V</td>
<td>296 sek</td>
</tr>
<tr>
<td>D&lt;sub&gt;BST&lt;/sub&gt;</td>
<td>DMA150E1600NA</td>
<td>SOT-227B (miniblock)</td>
<td>V&lt;sub&gt;RRM&lt;/sub&gt; = 1600V</td>
<td>163 sek</td>
</tr>
<tr>
<td>Alternative D&lt;sub&gt;BST&lt;/sub&gt;</td>
<td>SiC</td>
<td></td>
<td>V&lt;sub&gt;RRM&lt;/sub&gt; = 600V</td>
<td></td>
</tr>
<tr>
<td>R&lt;sub&gt;GATE&lt;/sub&gt;</td>
<td>±1%</td>
<td></td>
<td>(3.3Ω) depending on design (1/10W)</td>
<td></td>
</tr>
<tr>
<td>R&lt;sub&gt;GATE2&lt;/sub&gt;</td>
<td>±1%</td>
<td></td>
<td>10kΩ (1/10W)</td>
<td></td>
</tr>
<tr>
<td>R&lt;sub&gt;FB1&lt;/sub&gt;</td>
<td>±1%</td>
<td></td>
<td>1MΩ (400V)</td>
<td></td>
</tr>
<tr>
<td>R&lt;sub&gt;FB2&lt;/sub&gt;</td>
<td>±1%</td>
<td></td>
<td>13kΩ (1/10W)</td>
<td></td>
</tr>
<tr>
<td>C&lt;sub&gt;OUT&lt;/sub&gt;</td>
<td>ECE-P2WP152HX</td>
<td>Aluminum</td>
<td>450V (427V)</td>
<td>297 sek</td>
</tr>
<tr>
<td>R&lt;sub&gt;SENSE&lt;/sub&gt;</td>
<td>0.01Ω</td>
<td></td>
<td>(2.5W)</td>
<td></td>
</tr>
<tr>
<td>R&lt;sub&gt;SENSEfilter&lt;/sub&gt;</td>
<td>Chip resistor</td>
<td></td>
<td>220Ω (1/16W)</td>
<td></td>
</tr>
<tr>
<td>C&lt;sub&gt;SENSEfilter&lt;/sub&gt;</td>
<td>Ceramic X7R</td>
<td></td>
<td>1.5nF (100V)</td>
<td></td>
</tr>
<tr>
<td>C&lt;sub&gt;ICOMP&lt;/sub&gt;</td>
<td>Ceramic X7R</td>
<td>±10%</td>
<td>3300pF (50V)</td>
<td></td>
</tr>
<tr>
<td>R&lt;sub&gt;FREQ&lt;/sub&gt;</td>
<td>32.7kΩ</td>
<td></td>
<td>(65kHz)</td>
<td></td>
</tr>
<tr>
<td>C&lt;sub&gt;VCC&lt;/sub&gt;</td>
<td>Ceramic</td>
<td>±10%</td>
<td>0.1µF minimum</td>
<td></td>
</tr>
<tr>
<td>C&lt;sub&gt;VSENSE&lt;/sub&gt;</td>
<td>820pF</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>R&lt;sub&gt;CV&lt;/sub&gt;</td>
<td>Chip</td>
<td></td>
<td>16.2kΩ (1/10W)</td>
<td></td>
</tr>
<tr>
<td>C&lt;sub&gt;CV1&lt;/sub&gt;</td>
<td>Ceramic X5R</td>
<td>±10%</td>
<td>6.8µF (10V)</td>
<td></td>
</tr>
<tr>
<td>C&lt;sub&gt;CV2&lt;/sub&gt;</td>
<td>Ceramic X5R</td>
<td>±10%</td>
<td>0.47µF (10V)</td>
<td></td>
</tr>
<tr>
<td>PFC controller</td>
<td>UCC28180</td>
<td></td>
<td></td>
<td>22 sek</td>
</tr>
<tr>
<td>D&lt;sub&gt;START&lt;/sub&gt;</td>
<td>Switching</td>
<td></td>
<td>425V, 75A</td>
<td></td>
</tr>
<tr>
<td>D&lt;sub&gt;TURNOFF&lt;/sub&gt;</td>
<td>Schottky</td>
<td></td>
<td>40V, 2A</td>
<td></td>
</tr>
<tr>
<td>Sum:</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
4.4 Series-loaded resonant half bridge converter

The resonant half bridge was chosen mostly because the company already uses this topology in their other modulators with good results. The control and feedback system for two switches exists and can be implemented fast. A transformer ratio of 7.7:1 was chosen. The ratio should not be set to high because of the increased primary currents that will occur. It gives us a maximum average current at 10.3A. The $R_{DS(ON)}$ of the chosen MOSFETs is 14.5mΩ. $10.3A^2 \cdot 0.0145\Omega = 1.54W$ in conduction losses. The switching losses are considered low because of the soft switching with help of the resonant tank. The wanted transformer design was two C shape halves put together and the general power transformer low-mid frequency material P was chosen. The “right” core for the application became 0P45716UC with a fill factor of 44% which is within the recommended 60%. Even in the transformer it is more room for usage of litz wire instead of magnet wire.

According to the table the core losses should be 100mW/cm$^3$. The volume of the core is 27.9cm$^3$.

$$100\text{mW/cm}^3 \cdot 27.9\text{ cm}^3 = 2.79W$$

A way to archive the right leakage inductance is to make the transformer as good as it gets and then use a small air core inductor to increase it to the required 23.17µH. The difficulties with the transformer are a disadvantage of the resonant half bridge. If there was not any half bridge driver ready to use, a regular full bridge would be a good choice according to power level and power losses. A full bridge just has half the primary current because of full usage of the voltage drop to ground [1], although the driver timings can be difficult when two switch couples have to open and close simultaneously.

Table 5 shows the component list of the SLR half-bridge converter.

<table>
<thead>
<tr>
<th>Part:</th>
<th>Name:</th>
<th>Package:</th>
<th>Value:</th>
<th>Cost:</th>
</tr>
</thead>
<tbody>
<tr>
<td>2 x QSW</td>
<td>IXYS</td>
<td>IXFN210N30P3</td>
<td>$V_{DSS} = 300V$</td>
<td>2 x 245 sek</td>
</tr>
<tr>
<td>Transformer</td>
<td>2 x Mag-Inc.</td>
<td>0P45716UC</td>
<td>UR shape cores</td>
<td></td>
</tr>
<tr>
<td>2 x CICB</td>
<td></td>
<td></td>
<td>2 x 2.7µF (min 250V)</td>
<td></td>
</tr>
<tr>
<td>CRES</td>
<td></td>
<td></td>
<td>0.219µF (min 250V)</td>
<td></td>
</tr>
<tr>
<td>LLEAK</td>
<td></td>
<td></td>
<td>23.17µH leakage</td>
<td></td>
</tr>
<tr>
<td>Sum:</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Table 5: Component list for the SLR half-bridge converter
4.5 Second full-wave rectifier
The maximum RMS current reaches to 1.6A at 2kW and the forward voltage drop is then 0.9V. It gives a power loss of $2 \cdot 0.9V \cdot 1.6A = 2.88W$ total losses in the rectifier.

Component list in table 6.

<table>
<thead>
<tr>
<th>Part:</th>
<th>Name:</th>
<th>Package:</th>
<th>Value:</th>
<th>Cost:</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rectifier bridge 2</td>
<td>IXYS VBO40-16NO6</td>
<td>SOT-227B (miniblock)</td>
<td>$V_{RRM} = 1600V$</td>
<td>150 sek</td>
</tr>
<tr>
<td>Alternative bridge</td>
<td>SiC</td>
<td></td>
<td>$V_{RRM} = 1600V$</td>
<td></td>
</tr>
<tr>
<td>Sum:</td>
<td></td>
<td></td>
<td></td>
<td>150 sek</td>
</tr>
</tbody>
</table>

4.6 Energy efficiency results

Table 7: Energy efficiency results

<table>
<thead>
<tr>
<th>Losses (W):</th>
<th>100V RMS (1000W)</th>
<th>120V RMS (1500W)</th>
<th>230V RMS (2000W)</th>
</tr>
</thead>
<tbody>
<tr>
<td>First diode bridge</td>
<td>25,4</td>
<td>29,75</td>
<td>20,6</td>
</tr>
<tr>
<td>Bst inductor winding</td>
<td>5,59</td>
<td>7,66</td>
<td>3,68</td>
</tr>
<tr>
<td>Bst inductor core</td>
<td>1,74</td>
<td>1,74</td>
<td>1,74</td>
</tr>
<tr>
<td>Bst MOSFET</td>
<td>15,04</td>
<td>17,53</td>
<td>10,69</td>
</tr>
<tr>
<td>Bst diode</td>
<td>2,37</td>
<td>3,52</td>
<td>4,67</td>
</tr>
<tr>
<td>Bst Rsense</td>
<td>1,64</td>
<td>2,26</td>
<td>1,08</td>
</tr>
<tr>
<td>HB 2 x MOSFET</td>
<td>17,5</td>
<td>17,5</td>
<td>17,5</td>
</tr>
<tr>
<td>HB transformer winding</td>
<td>0,5</td>
<td>1</td>
<td>2</td>
</tr>
<tr>
<td>HB transformer core</td>
<td>2,79</td>
<td>2,79</td>
<td>2,79</td>
</tr>
<tr>
<td>Second diode bridge</td>
<td>1,44</td>
<td>2,16</td>
<td>2,88</td>
</tr>
<tr>
<td>Total</td>
<td>74,01</td>
<td>85,91</td>
<td>67,63</td>
</tr>
<tr>
<td>Efficiency</td>
<td>93%</td>
<td>94%</td>
<td>96%</td>
</tr>
</tbody>
</table>

Table 7 shows max load, worst case energy efficiency at three different part of the world. The goal was to reach an efficiency of 95%. This design and components get very close by having really good switching diodes with close to zero reversion recovery like the Silicon Carbide (SiC) type. They are more expensive but will manage the work as boost diode and second half bridge better. Two parts are marked in the component list and it might be recommended to look at alternatives for these. Why it was not chosen from the beginning was a wish to use the power supply steel box as a heat sink, then SOT-227B is a good package and the fact that no SiC components with SOT-227B package was found. They usually use some kind of TO-220 similarities and that is why two different designs in SolidWorks was made (figure 18 and 19).
4.7 Design examples

Figure 18: Design example with SOT-227B (miniblock) package

Figure 19: Design example with TO-220 package and associated heat sinks
Figure 18 and 19 show two different designs that are made in SolidWorks with respect to the size of the current power supply (341mm × 135mm × 88mm) and “as close as possible” size of the components in the component lists. According to figure 18 and 19 all the components fit but this needs to be confirmed with a proper PCB design.
5 Conclusions
The outcome of this project is a single phase power supply design with PFC that can be used all over the world and that meets the EN61000-3-2 standard of current harmonics. The voltage output will be ~1500V. The conducted EMI disturbances that can occur are removed with a second order low pass filter to meet the EN55011 standard. A physical design is made in SolidWorks on present power supply’s size and according to the size of the proposed components in the component list. The design works according to required space. An energy efficiency calculation is made by datasheet values and equations and meets the aim of 95% energy efficiency in most cases. A test built is necessary before production.
6 References

[1] Abdel-Rahman, S. (2012, September). Retrieved from http://www.infineon.com/dgdl/Application_Note_Resonant+LLC+Converter+Operation+and+Design_Infineon.pdf?folderId=db3a30431a5e32f2011a77f9c03e6cb4&fileId=db3a30433a047ba0013a4a60e3be64a1


