MIDDLE EAST TECHNICAL UNIVERSITY

ELECTRICAL & ELECTRONICS ENGINEERING

EE464 - STATIC POWER CONVERSION II
Hardware Project - Forward Converter #1
PowerLovers
Simulation Report

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23.03.2020

Contents

1 Introduction

Within the scope of the "EE464 - Static Power Conversion II" class offered at METU EEE in Spring'20, a hardware project is to be completed and this report is written in order to inform the reader about the selected topology, the conceptual design of the project and the simulations results. In the first part of this report, different topologies of regulated power supplies, namely the flyback and forward converters, are compared. Their advantages and disadvantages are stated. In the second part, the selected topology is announced and reasoned why it is chosen so. An important part of designing regulated power supplies is the magnetic design, therefore a separate part is devoted for the magnetic design of the transformer in the topology. Simulation results under ideal conditions are presented in the following subsection and other components are selected considering these simulations. Lastly, non-idealities introduced by the selected components and how to deal with them are discussed. The group's main goal in this project is to obtain a robust, simple solution to a regulated power supply with pre-specified criteria. Some of the bonuses are aimed for and they are also discussed in detail in the following subsections. Our group is highly motivated and well-prepared to overcome the difficulties on the way and hopefully will present a satisfying result at the end of this project.

2 Problem Definition

In this hardware implementation project, there are two available isolated power supply topologies : Flyback and forward converter topologies.

Three different set of specifications for each topology are presented, adding up to a total of six options for the project. However, they are not that different from each other. Two of the topologies require rectification due to the AC Power input and the remaining four only have small differences. Output power levels range from 48W to 60W, ripple requirements are restricted to 2%.

The requirements for all options are not given here in detail. Only the ones for the chosen topology are presented in detail in the following subsections.

3 A Comparison of Two Topologies

Before the chosen topology is announced and conceptual design is examined in detail, a brief description for each topology is presented here to better understand both of the options. After that, a comparison between the two solidifies the reasoning for the selected topology.

3.1 Flyback Converter

The flyback converter is actually derived from the general buck-boost topology. Main energy transferring component was inductor in the buck boost topology. In flyback converter, to ensure isolation between input and output stages, that inductor is replaced by a transformer and the primary winding of the transformer is the energy transfer component.
The flyback converter topology is given in Figure 3.1.1.

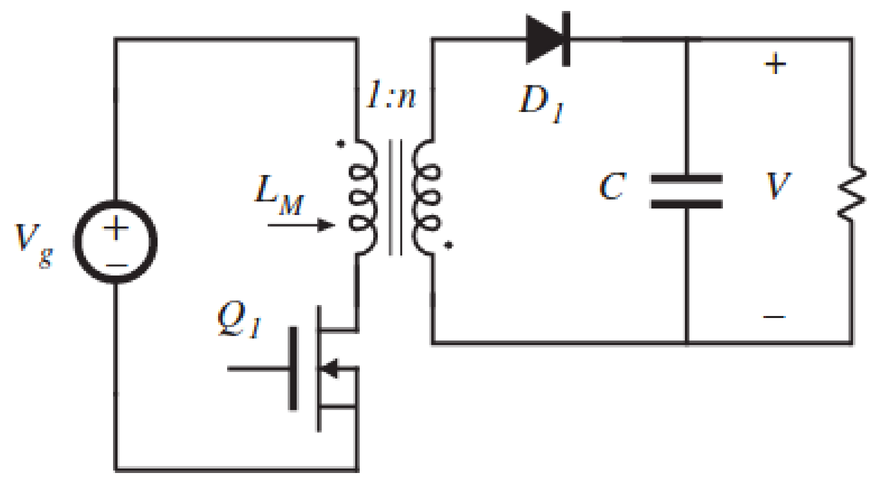

Figure 3.1.1: Flyback Converter Topology

During the ON state of the switch, primary winding of the transformer is connected directly to the source. Magnetizing branch is charged during this period. Due to the dot convention of the transformer windings, secondary winding diode is reverse biased and no current is sent to that side. Load is fed by the output capacitor.

During the OFF state, magnetic flux stored in the transformer is utilized. Magnetizing branch discharges through the primary winding and diode in the secondary side is now conducting. Primary winding current and flux stored inside the core of the transformer are diminishing.

It can be seen that L_m component of the transformer is of utmost importance to understand the operation of this topology. Writing the voltage-seconds rule for this component also yields the transfer function for the converter. During the on state, $V_{L_m} = V_d$ and during the off state, $V_{L_m} = \frac{-N_1}{N_2} V_o$. The inductor voltage at steady state should equal to 0.This gives the following transfer function for the flyback converter as follows.

$$\frac{V_o}{V_d} = \frac{N_2}{N_1} \frac{D}{1-D} \tag{1}$$

Added fraction by the transformer turns ratio broadens the operation range compared to a typical buck-boost converter.

3.2 Forward Converter

Second available option is the forward converter. It is derived from a previous topology, namely buck converter, by introducing a transformer in between input and output stages in order to achieve galvanic isolation of the two.

In Figure 3.2.1, the forward converter topology can be observed.

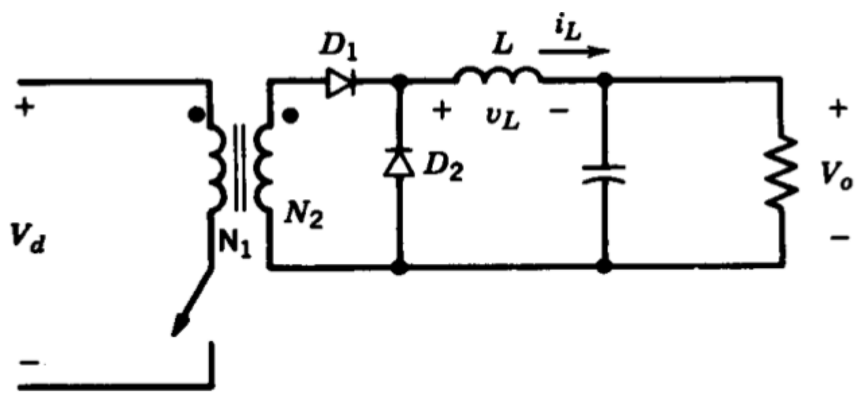

Figure 3.2.1: Forward Converter Topology

Assume the transformer is ideal for now to obtain the transfer function of the topology.
During the ON state of the switch, D_1 is conducting and D_2 is reverse biased. Therefore, the inductor voltage can be written down as

$$V_L = \frac{N_2}{N_1} V_d - V_o$$

Let us examine now the OFF state of the switch. D_1 is now reverse biased on D_2 forms a free-wheeling path for the inductor. Load is fed by the stored energy in the inductor and the capacitor.

Inductor voltage is easy to obtain and is as follows.

$$V_L = -V_o$$

Discussion so far end with applying the voltage-seconds rule for the inductor and this yields the transfer function of the forward converter topology as

$$\frac{V_o}{V_d} = \frac{N_2}{N_1} D \tag{2}$$

However, a question might arise rightfully at this point: how does this topology deal with the magnetizing current? During the OFF state, L_m has no path to discharge and it seems to be charged again and again during ON state. This situation ends up saturating the core and threatens the proper operation of the converter.

A variety of solutions for this problem are available. A snubber circuitry connected in parallel to the primary winding, a two switch topology are among them. A more practical and wide-spread application ,however, is adding a third reset winding. The resultant topology can be seen in Figure 3.2.2.

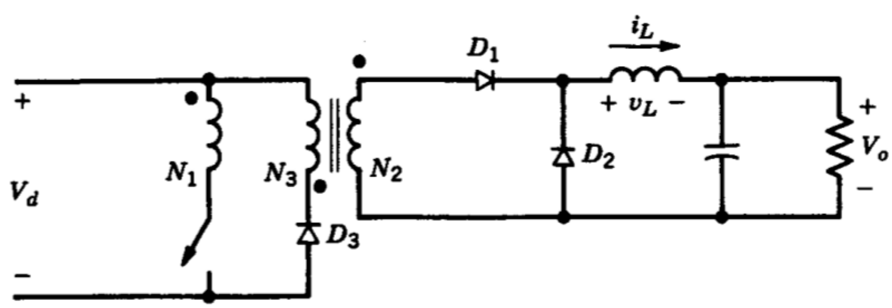

Figure 3.2.2: Practical Forward Converter Topology

3.3 Flyback Converter: Summary

- Advantages

  Simple topology with small component count. More economical in most of the times.

- Drawbacks

  Requires a gapped core most tof the times due to the energy storage in transformer.

  Without a filter, both input and outppput currents are discontinuous.

3.4 Forward Converter: Summary

- Advantages

  Magnetic design is simpler, no gap is required.

  Output current is filtered by the nature of the topology.

  Direct power transfer, a better utilization of transformer.

- Drawbacks

  More components are required. Complexity and cost are increased.

  Due to the reset winding, voltage stress on switching element is higher.

4 Topology Selection and Reasoning

In the second part of this report, two available topologies are discussed and compared. Regarding the advantages and disadvantages of each possible solution, we, as PowerLovers, choose to move on with the Forward Converter topology, to be more exact, #FOR1 from the offered specifications set.

We gave this decision because of the simpler magnetic design and more stable output current inherited in the topology itself. Although it is a bit more complex to work with, team is motivated to yield a fruitful output -a properly working, well engineered project- and ready to embrace a more challenging overall design.

5 Design of the Forward Converter

The most important element in isolated power supplies is the transformer. Without knowing the transformer properties, it is not possible to move on with the simulations. Hence, the design process is now orderly. Output inductor and capacitor can also be chosen by theoretical knowledge.

Although it is possible to have a "pre-simulation" idea on the voltage stresses or current carrying capabilities for the semiconductor devices in the topology, it is safer to choose them in the guidance of the simulation results. Some of the non-idealities will also be involved in the simulations as well, allowing us to make more accurate assumptions about component selection.

5.1 System Level Design

In this part, we need to look at general specifications of the forward converter. Our customer, from Habelsan, asked us to satisfy these specifications:

- Input voltage rate: $24V - 48V$
- Output voltage: $15V$
- Output power: $48W$
- Output voltage ripple: 2%, maximum
- Line and Load regulations: 2%, maximum

Firstly, we need to look at input-output voltage relationship of the forward converter. Ratio between the two is derived in the previous parts as:

$$\frac{V_o}{V_d} = \frac{N_2}{N_1}D \tag{3}$$

Output current's average value is $\frac{48W}{15V} = 3.2A$. As mentioned before, demagnetization of the core in the forward converter is important. Among different techniques,it is decided to use a reset winding. Turn number of the reset winding is chosen to be equal to that of the primary winding, conventionally and we will stick to that convention as well. This situation puts a restriction on the duty cycle to ensure proper demagnetization of the core.Duty cycle for the operation should be $D < 0.5$.

Turns Ratio & Duty Cycle

In the project, required voltage transfer ratios are 48V to 15V, and 24V to 15V. It is important to take into account that the limit of duty cycle is 0.5, and it is very crucial to reset all the core.

Using the input output equation and duty cycle restrictions, we choose the turn ratio between first and secondary winding as:

$$\frac{N_2}{N_1} = \frac{3}{2}$$

Using these ratio, it is now easy to find required duty ratios. For 48V input,

$$D = \frac{48}{15} = \frac{3}{2} \implies D_{min} = 0.209$$

and For 24V input,

$$D = \frac{24}{15} = \frac{3}{2} \implies D_{max} = 0.4167$$

As can be seen, the duty cycle ratios obey the restrictions and they also allow a margin to compensate for the non-idealities.

Frequency

As switching frequency of the MOSFET, it is important to note that higher frequencies increases the switching losses. Also, at high frequencies, skin depth of the cable decreases dramatically. Moreover, it is also needed to be pointed that at higher frequencies hysteresis losses increase in the core. Therefore, we decided to keep our frequency less than 30kHz, $f < 30kHz$

Secondly, as frequency decreases, the amount of ripple at the output increases due to longer switching periods. Furthermore, it is important to keep the frequency at inaudible range so that the converter is not noisy. We decided that frequency should be higher than 20kHz, $f > 20kHz$.

Combining these two

$$20kHz < f < 30kHz$$

In the simulations, we decided as the best frequency would be 25kHz. Therefore, for our forward converter the frequency is $f = 25kHz$

5.2 Transformer Design

The ratio between the primary and secondary windings is determined and the reset winding turn number will be equal to the primary winding turn number. While deciding on the specific numbers, core parameters are important. The first of the restrictions come from the renowned Faraday's law of induction,

$$e = -N\frac{d\phi}{dt}$$

where e is the induced voltage and ϕ is the magnetic flux.

Let us change this formula a bit to set our restrictions. We know $\phi = B.A$. Here A will be A_e, cross section area, and B will be B_{sat} to account for the extreme cases. Maximum values for other parameters will be used as well to make sure that is not saturating. The following formula is obtained for the primary winding.

$$N_1 > \frac{V_{d,max}D_{max}T_s}{B_{sat}A_e} \tag{4}$$

Core should be selected at this point. Size of it is important to fit the windings in it. The greater the core is, the more capable it is to carry flux. But it should also be practical to use, not so bulky if possible. Two candidates at this point are the E-cores by *Magnetics Inc.*, models **00K6527E060** and **00K7228E060**. Both made up of "Kool Mμ" material, and have saturation densities as $B_{sat} = 1\,T$. Cross section area for the first one is $A_e = 540\,mm^2$.

Inserting the parameters from our case,

$$N_1 > \frac{48V * 0.5 * 40\mu s}{1T * 540mm^2} = 4.05 \tag{5}$$

Actually, that number is really small. It results in a very low magnetizing branch inductance and causes really high magnetizing and demagnetizing currents. This is not something desired. Choosing the number larger than what is found, also keeping the window area of the core in mind, is reasonable.

The turns number for windings are determined as

- Primary winding, $n_1 =$ **24 turns**
- Reset winding, $n_3 =$ **24 turns**
- Secondary winding, $n_2 =$ **36 turns**

5.3 Cable Selection

In the transformer design, we concluded that we are going to use an E core with window area of $530mm^2$, and in practical cases maximum fill factor achievable is stated at 50%. We need to take these two parameters into account.

Moreover, it is important to notice that, the input and output current RMS is 4 Amperes, so it is very important for cable to be able to carry 4A RMS in 20-30kHz frequency range.

Between several options, we concluded the cable AWG14. After the practical tests, the cable may be changed.

Parameters of the cable at decided frequencies for 1 meter of copper wire:

For 20kHz **AC Resistance = 0.00953, Skin depth = 461μm**,

For 30kHz **AC Resistance = 0.01085, Skin depth = 376μm**

Other possible selections are: AWG13, AWG14, AWG15, AWG16

5.4 Capacitor Design

For a DC-DC converter, it is very crucial to have ripple free output voltage in order to maintain a stable operation. Therefore, we need to introduce a capacitor to the output so that output ripple is less than 2%.

Using following:

$$\frac{\Delta Q}{C} = \Delta V$$

$$\frac{I_o(1-D)T_s}{\Delta V} < C$$

$$\frac{3.2A * 0.77}{25kHz * 15V * 0.02} < C$$

$$328\mu F < C$$

Also, it is important to notice that rated voltage of our capacitor has to be higher than 15V

$$15V < V_{c,rated}$$

5.5 Inductor Design

It is important to have less ripple on the output current, and in the average, we can say that inductor current is equal to the output current, $I_{L,avg} = I_o$. Therefore it is important to have a ripple free output. We decided to have a ripple less than 10% on the inductor.

$$\Delta I = \frac{1}{L}\int_0^{DT_s} V_L dt < 0.32A$$

$$\Delta I = \frac{1}{L}\int_0^{DT_s} (V_s\frac{N_2}{N_1} - V_o)dt < 0.32A$$

$$1.3mH < L$$

It is important to have an inductor with an inductor higher than 1.3mH.

5.6 Switch Design

In these voltage levels and frequency levels it is proper to use MOSFET due to their fast recovery. It is also easy to implement a MOSFET into a converter. While designing the MOSFET, it is very important not to exceed its rated voltage and current values.

In the simulations with idealities, we came up with the values that:

- Switches' stress have their maximum values for input of 48V naturally. We can say that switches must be endurable minimum of 100V rated reverse voltage and 10A forward voltage.
- Switches must be proper to operate at 25kHz range, reverse recovery times should be appropriate.

The results of the simulations will be presented in the next section, in the Figure 6.1.5 and 6.2.5

5.7 Diode Design

Silicon diodes are proper because they are cheap in cost and they have proper operation for this implementation range.

In the simulations with idealities, we came up with the values that:

- Diode' stress have their maximum values for input of 48V naturally. We can say that switches must be endurable minimum of 100V rated reverse voltage and 4A forward voltage.
- Diode must be proper to operate at 25kHz range, reverse recovery times should be appropriate.

Results of simulations will be introduced at next part, in Figure 6.1.6, 6.2.6, 6.2.7, 6.1.7

5.8 Feedback & Switch Driver Circuitry

To keep the voltage constant at 15V and to satisfy the line/load regulations, an analog controller IC is to be used. A very popular PWM controller IC for power supplies, **TL494** by *Texas Instruments* will be our choice. Most important considerations for choosing a control circuitry is built in in TL494. An internal 5V reference, adjustable frequency oscillator, soft starting capability and dead-time control (although not used in our application) are some of the merits of TL494.

A resistive voltage divider branch connected to the output is feeding the error amplifier together with a similarly modified branch from the internal reference voltage pin.

TL494 is able to drive two switches together, but only one of the outputs is to be used in our project. Soft starting and current limiting features are to be adapted to our application, though. Additional circuitry for them can be found in the application notes regarding TL494.

To ensure electrical isolation between input and output sections of the converter, an optocoupler is to be used to drive the switching element. Our choice for this element is **TLP250** by *Toshiba*. It is a familiar IC for us, since it is used previously on the hardware project of EE463.

6 Simulations

Below in the Figure 6.0.1, the simulation model schematic can be seen.

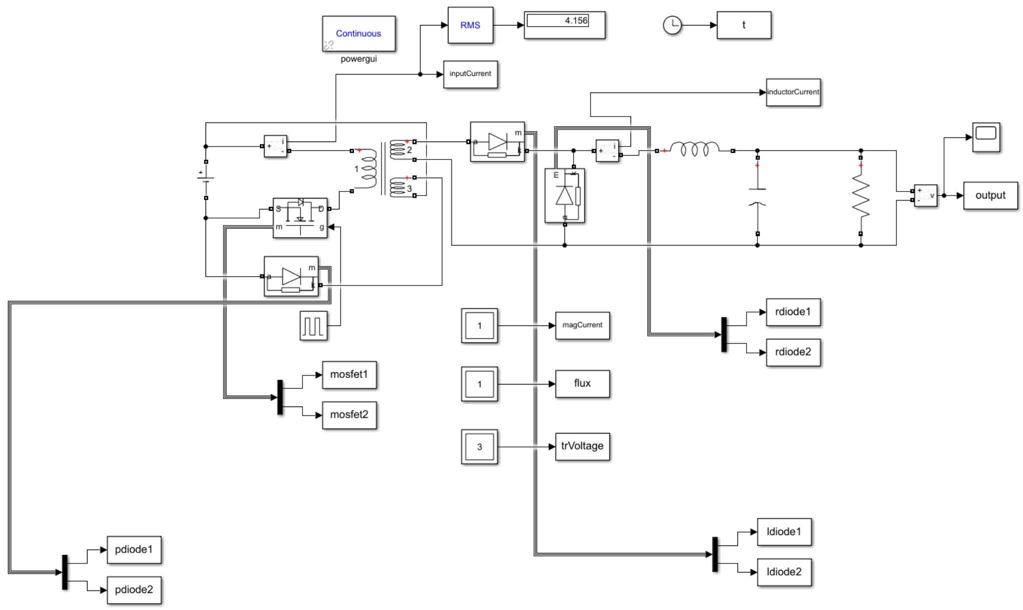

Figure 6.0.1: Simulation model

6.1 Simulations at 24V Input

6.1.1 Output Simulation

Output of the forward converter can be seen in the Figure 6.1.1. Ripple is less than 2%, and output is fixed at 15V, at duty cycle 0.42.

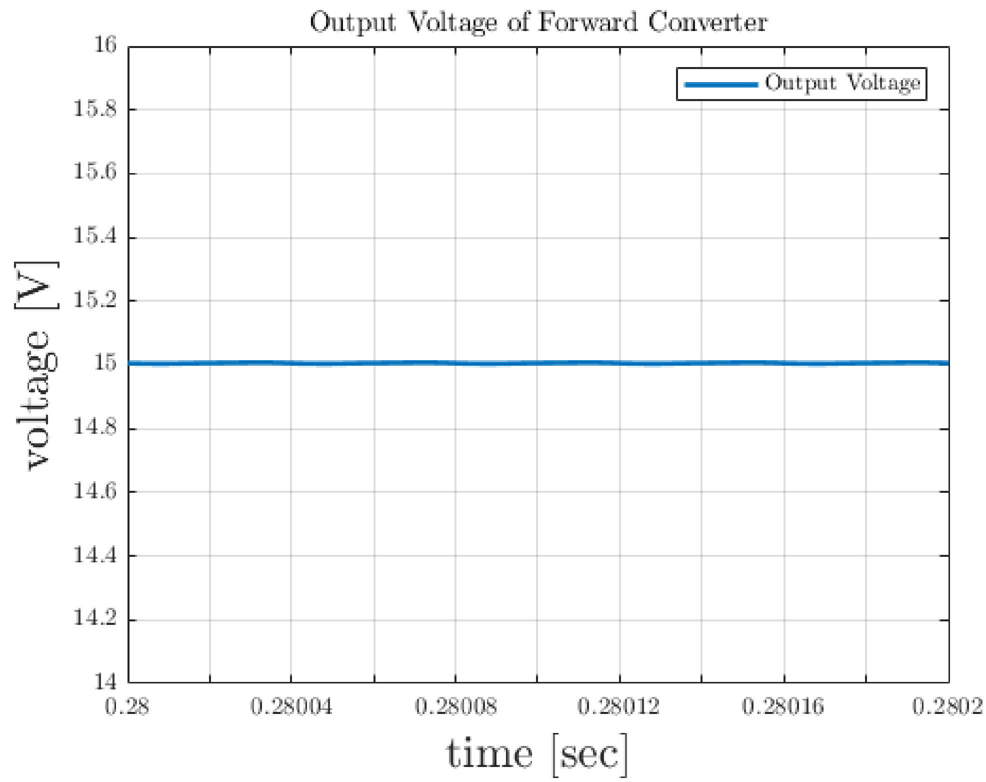

Figure 6.1.1: Output waveforms

6.1.2 Input and Transformer Simulation

Input current of the forward converter can be seen in the Figure 6.1.2. Average current is around 4A.

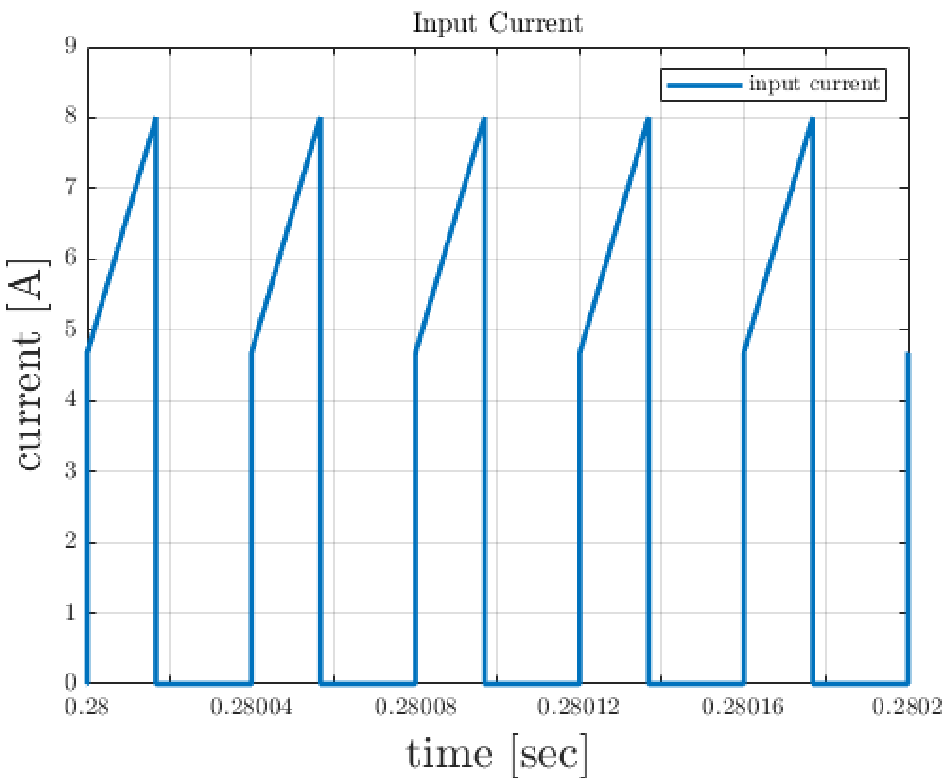

Figure 6.1.2: Input current waveforms

Magnetizing current of the forward converter can be seen in the Figure 6.1.3. Average current is around 2A.

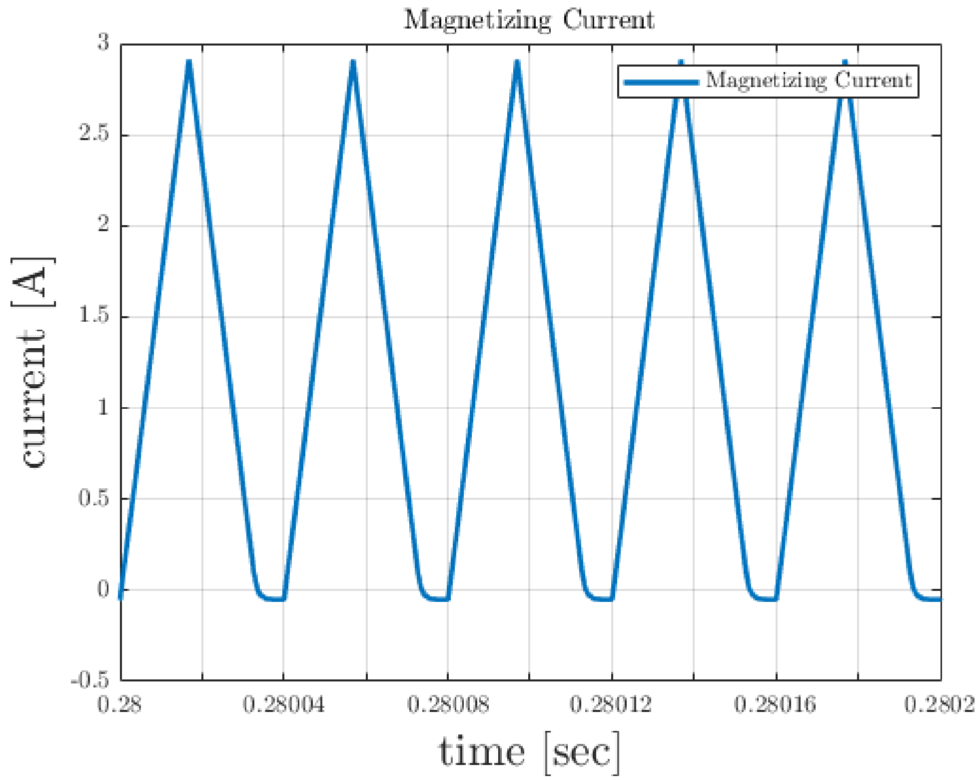

Figure 6.1.3: Magnetizing current waveforms

6.1.3 Inductor Simulation

Inductor current of the forward converter can be seen in the Figure 6.1.4. Average current is around 3.2A, and ripple is around 300mA as expected.

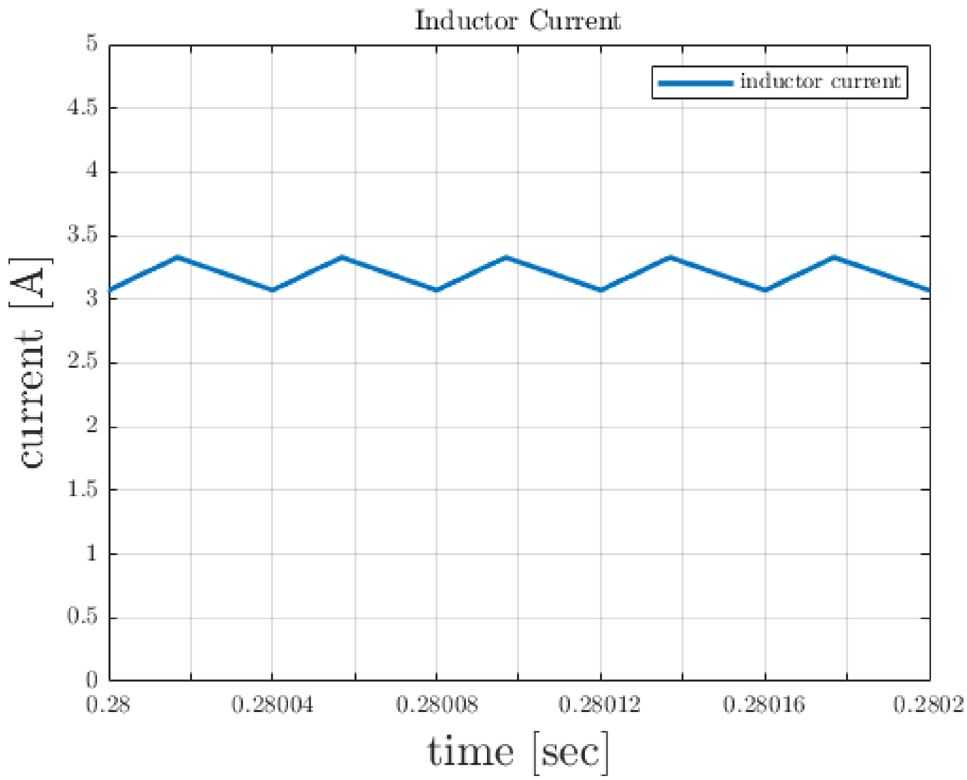

Figure 6.1.4: Inductor current waveforms

6.1.4 Switch Simulations

MOSFET

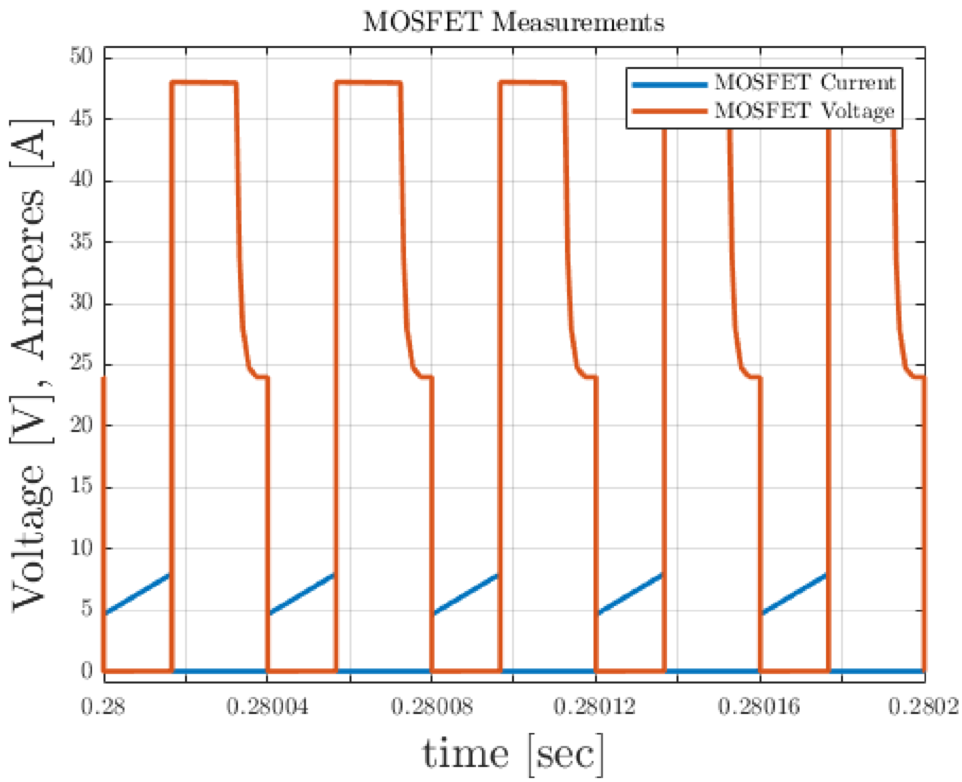

Figure 6.1.5: MOSFET stress waveforms

As we see, MOSFET current maximum is 8A, where maximum voltage is 48V.

6.1.5 Diode Simulations

In the simulations concerning diode parameters and limitations, we have following results:

Primary Side Diode

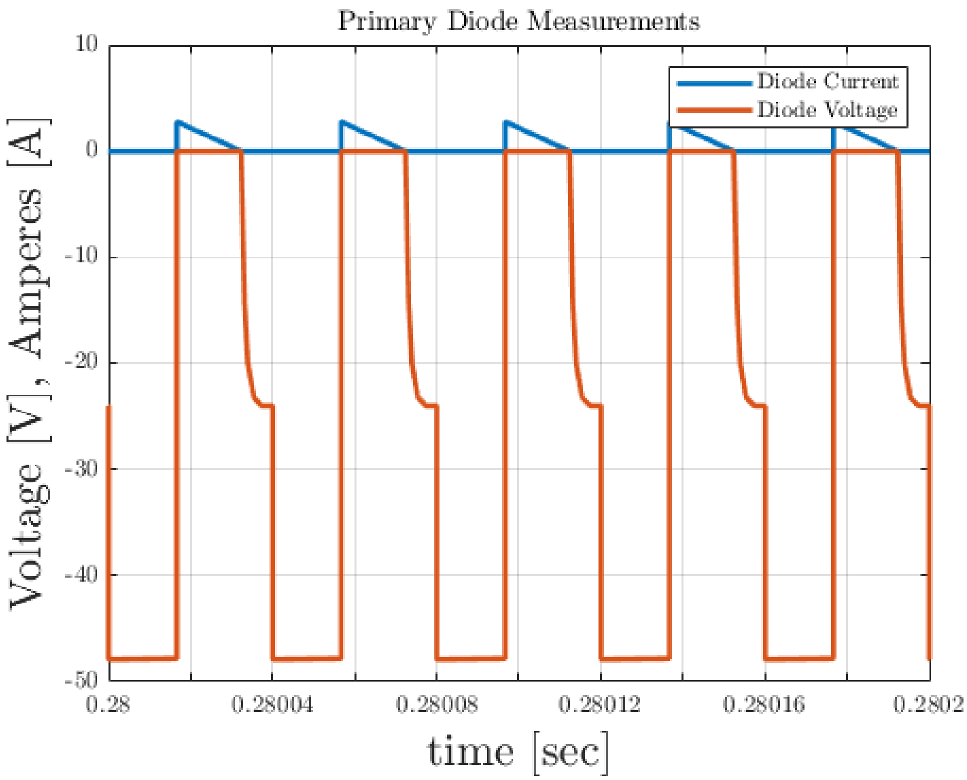

Figure 6.1.6: Primary diode stress waveforms

As we see from the Figure 6.1.6, the maximum current on the diode is 2.75A, where maximum reverse voltage is -48V. These parameters are important while selecting components.

Secondary Side Diode 1

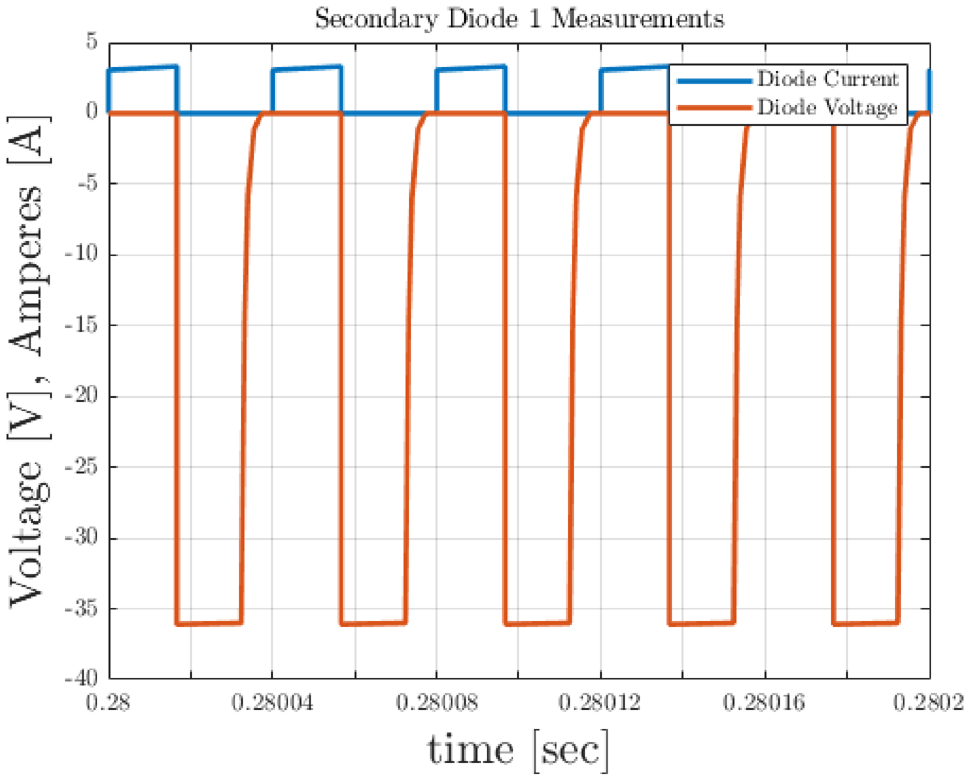

Figure 6.1.7: Secondary side diode 1 stress waveforms

As we see from the Figure 6.1.7, the maximum current on the diode is 3.32A, where maximum reverse voltage is -36V. These parameters are important while selecting components.

Secondary Side Diode 2

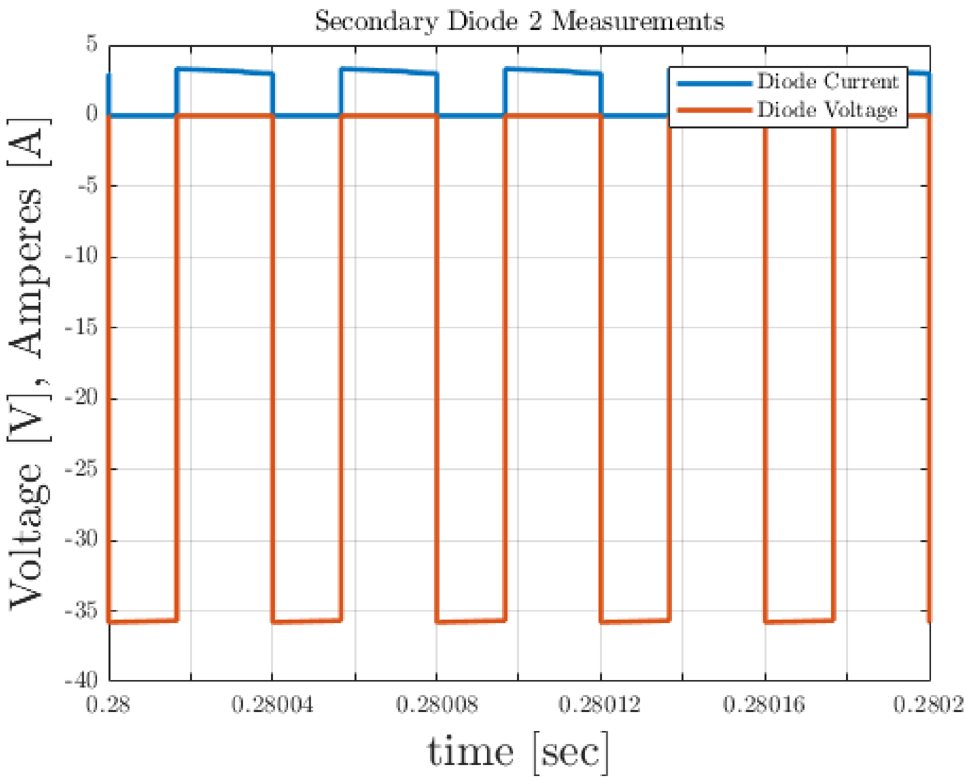

Figure 6.1.8: Secondary side diode 2 stress waveforms

As we see from the Figure 6.1.8, the maximum current on the diode is 3.32A, where maximum reverse voltage is -36V. These parameters are important while selecting components.

6.2 Simulations at 48V Input

6.2.1 Output Simulation

Output of the forward converter can be seen in the Figure 6.2.1. Ripple is less than 2%, and output is fixed at 15V, at duty cycle 0.209

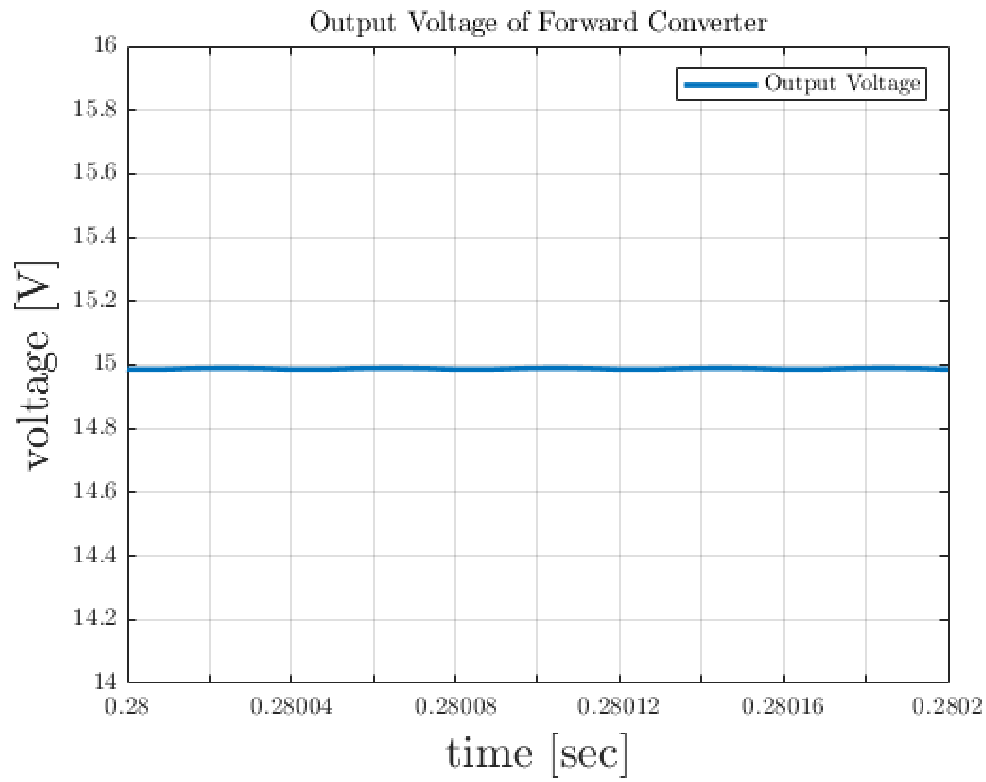

Figure 6.2.1: Output waveforms

6.2.2 Input and Transformer Simulation

Input current of the forward converter can be seen in the Figure 6.2.2. Average current is around 4A.

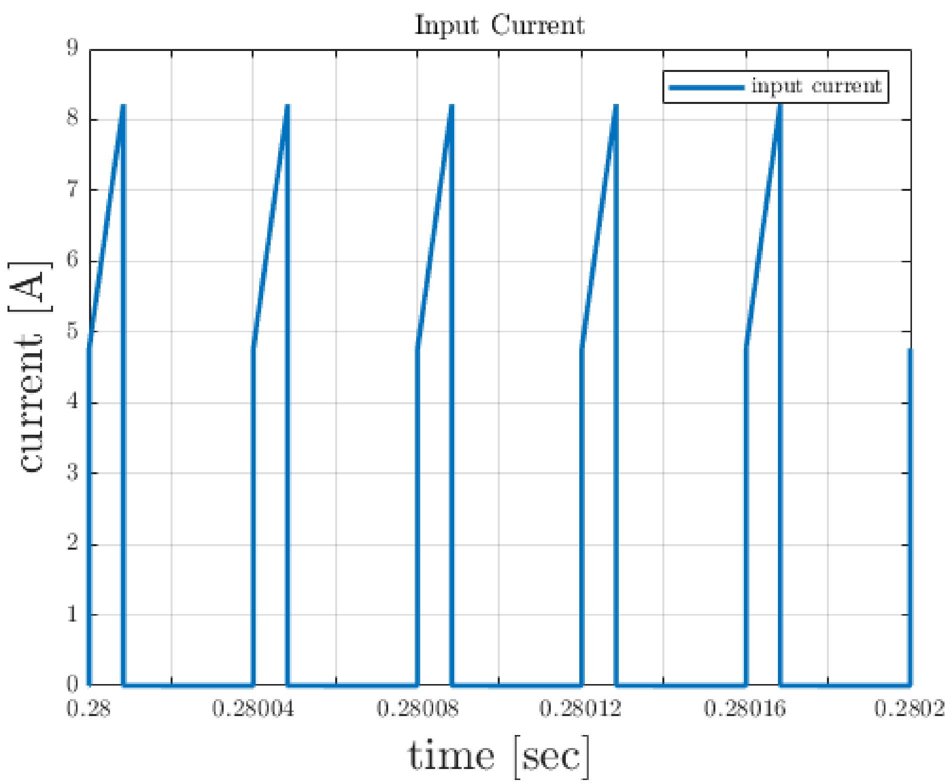

Figure 6.2.2: Input current waveforms

Magnetizing current of the forward converter can be seen in the Figure 6.2.3. Average current is around 2A.

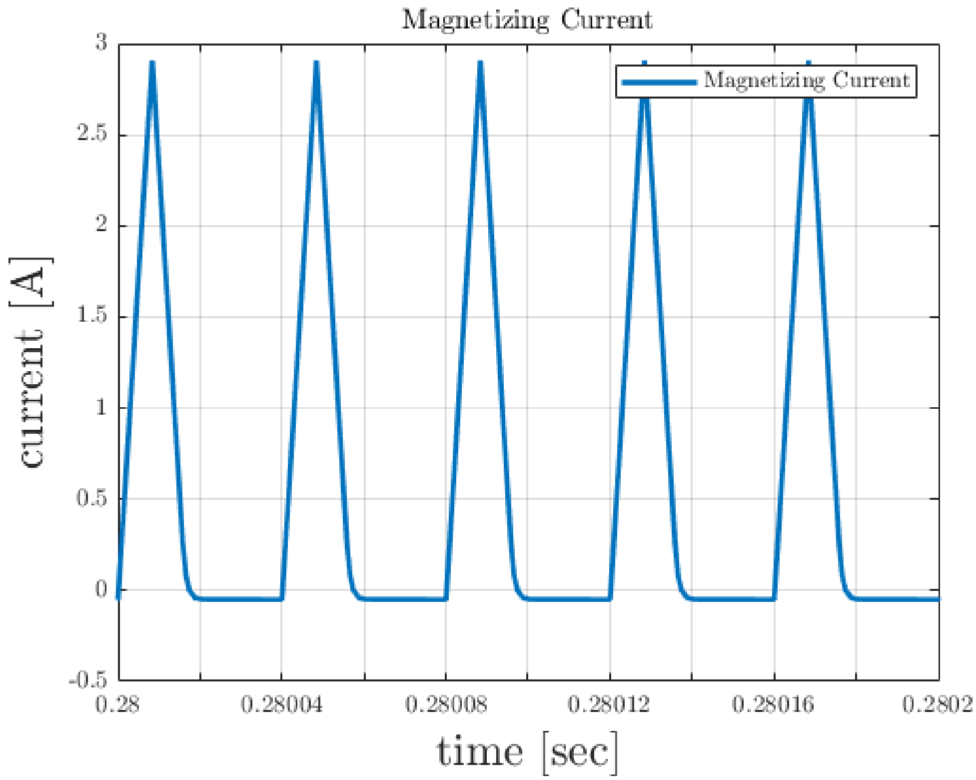

Figure 6.2.3: Magnetizing current waveforms

6.2.3 Inductor Simulation

Inductor current of the forward converter can be seen in the Figure 6.2.4. Average current is around 3.2A, and ripple is around 300mA as expected.

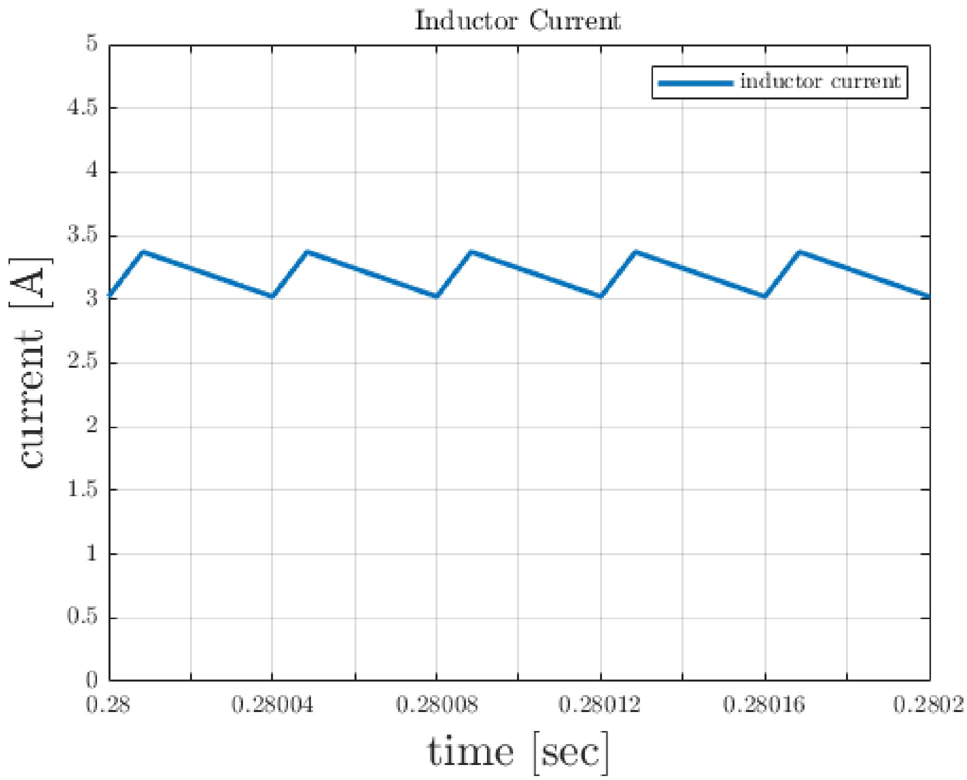

Figure 6.2.4: Inductor current waveforms

6.2.4 Switch Simulations

MOSFET

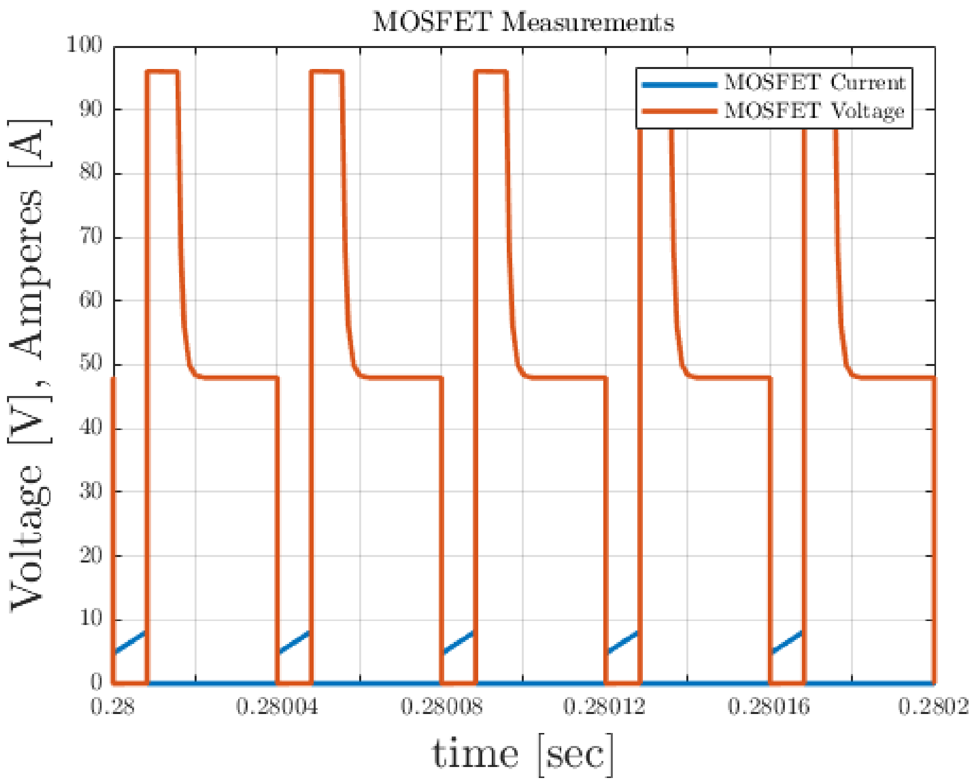

Figure 6.2.5: MOSFET stress waveforms

As we see in Figure 6.2.5, MOSFET current maximum is 8A, where maximum voltage is 96V.

6.2.5 Diode Simulations

In the simulations concerning diode parameters and limitations, we have following results:

Primary Side Diode

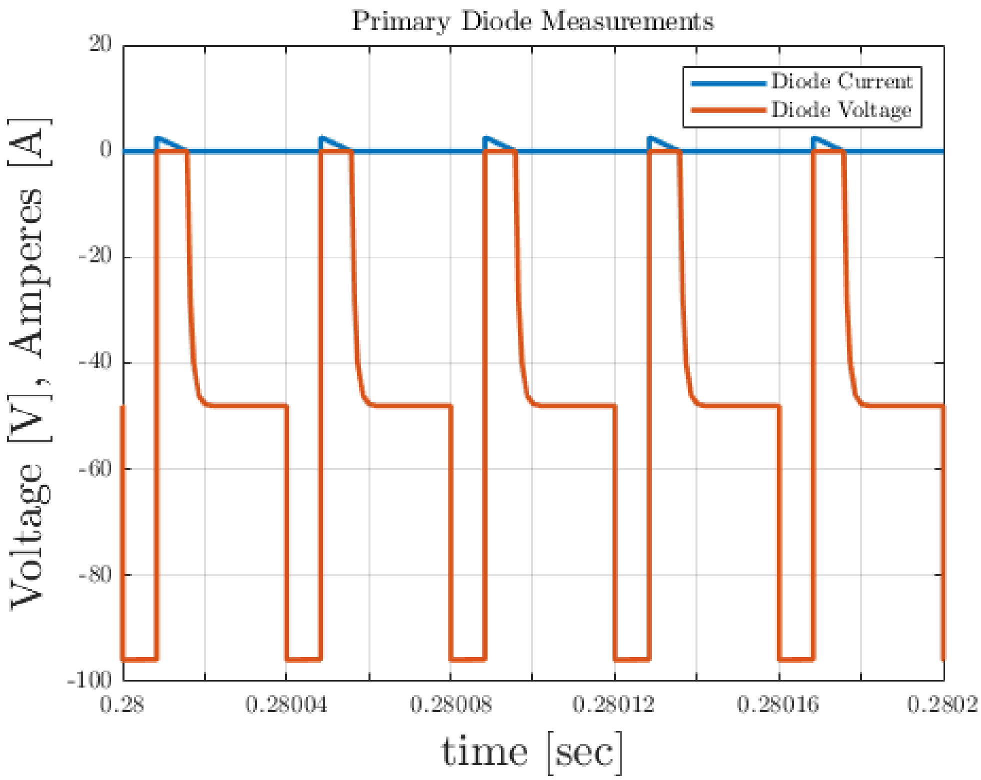

Figure 6.2.6: Primary diode stress waveforms

As we see from the Figure 6.2.6, the maximum current on the diode is 2.6A, where maximum reverse voltage is -95V. These parameters are important while selecting components.

Secondary Side Diode 1

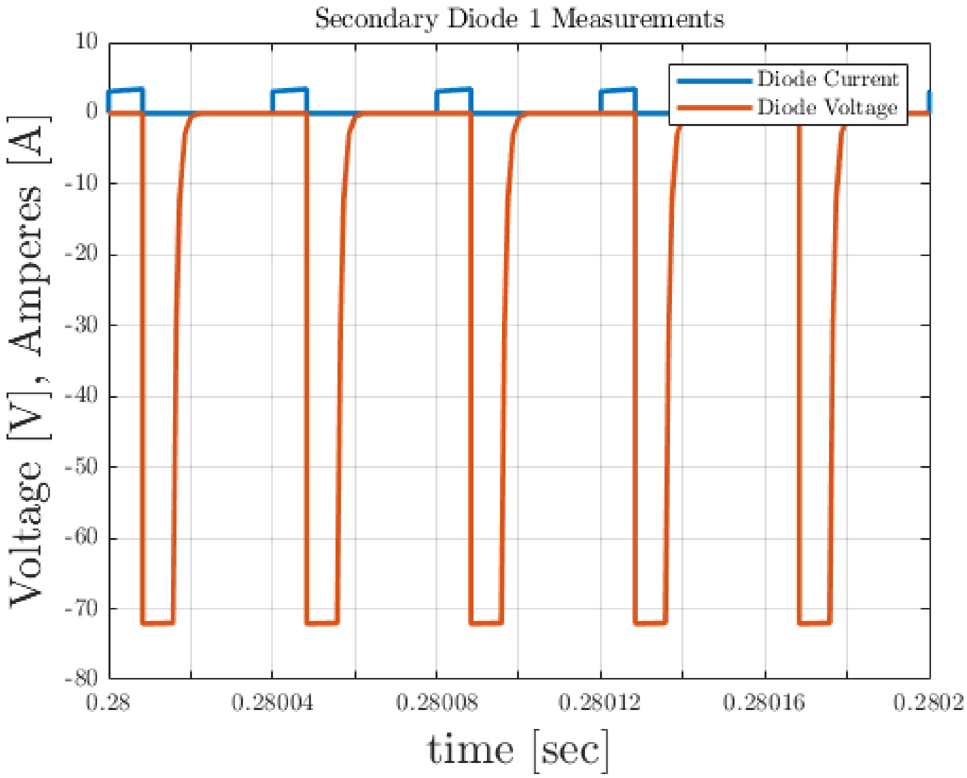

Figure 6.2.7: Secondary side diode 1 stress waveforms

As we see from the Figure 6.2.7, the maximum current on the diode is 3.5A, where maximum reverse voltage is -72V. These parameters are important while selecting components.

Secondary Side Diode 2

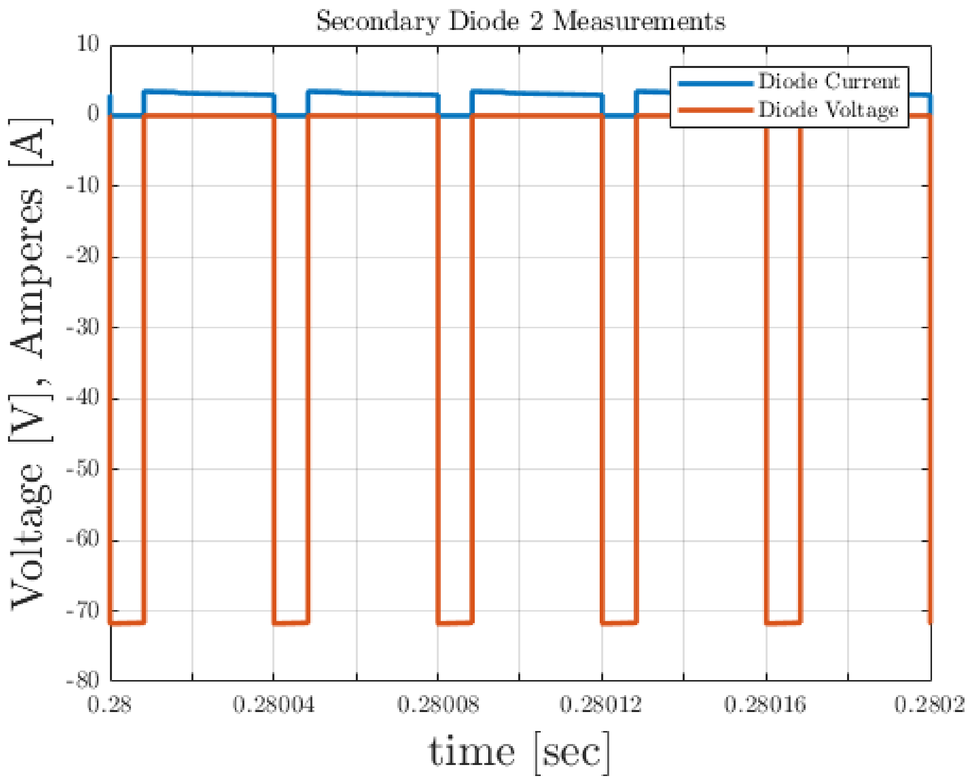

Figure 6.2.8: Secondary side diode 2 stress waveforms

As we see from the Figure 6.2.8, the maximum current on the diode is 3.5A, where maximum reverse voltage is -72V. These parameters are important while selecting components.

7 Component Selection

In the previous part, design and simulations are presented. From the simulation results, device restrictions are also obtained. We can now decide on the components. As a team, we have tried to restrict ourselves to the available component list.

Switching Device

Since we are not working with high level voltages, a MOSFET can be used safely. At maximum input voltage, 48 V, MOSFET drain-to-source is voltage is found to be 96 V. Peak current value that it should withstand is 8A, as can be seen from 6.2.5.

A proper choice can be **IRF740N**. ıt can withstand 10A under 400V. Datasheet can be found here.

Diodes

At the maximum input voltage case, 48V, voltage stress on the primary side diode is found to be 95V and for the secondary sides it is 72V. Lower input voltage case results in lower stresses. Currents they carry are varying, but the maximum value is 3.5A.

Since we try to restrict our components to the available components list, **MUR1560G** seems a suitable choice. It can withstand 600V and can carry 15A. Even these values are more than what we need. Datasheet can be found here.

Output Capacitor

Restrictions for the output capacitor were found as minimum 15V voltage handling capability and at least 328 μF of capacitance. In the available components list, there is no such component.

An alternative is found from Digikey. **KHD250E477M99C0B00** model by *THD Series* is fitting for our needs. It can withstand 25V and has 470 μF of capacitance. Datasheet can be found here

Another alternative is to use paralleled capacitors from the list. 3-4 of them will do the job.

Output Inductor

The inductor for the output side is to be wound by the team. A toroidal core can be used. A 2 mH inductor is planned to be on the safe side regarding the current ripple.

Cable should be able carry 3.2A average current, continuously. Let us choose an **AWG16** cable for the winding. It has a cross sectional area of $1.31mm^2$.

For the core, **0077442A7** model toroidal core by *Magnetics Inc.* is a good choice. A_l value for this core is $202nH/T^2$ and to get 2mH of inductance, it requires **100 turns**. With a fill factor of 0.5, the required area is around $260mm^2$, considering an AWG16 standard cable is used. Window area of the core is $427mm^2$. It means that winding that cable 100 turns is possible with a safety margin.

8 Conclusion

In this project, our goal is to design a DC to DC converter. Our team has chosen forward converter configuration, #FOR1, for the project. In this report, we introduced possible topologies and the rationale behind our selection. Then, we theoretically calculated the needs of the project and the components. A detailed component by component calculation has been provided. Moreover, simulations for each component have been done and introduced. With the light of the simulations, our selections and theoretical calculations have been supported. It is now possible to say that the design introduced in this simulation report is doable, and properly functional. In conclusion, EE464 lectures have prepared us academically. We strongly believe that this project will enhance our understanding, and provide broader vision into power electronics area.