DESIGN ZERO-VOLTAGE SWITCHING DC-DC BUCK CONVERTER

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This report proposes an integrated, high switching frequency, zero-voltage-switching dc-dc buck converter for battery charger application. The design and analysis of dc-dc buck converter with integrated inductor is presented. The converter has been optimized to convert 12V input voltage to 5V at 1.5A maximum load current at 50MHz switching frequency. The converter has been simulated using an ORCAD 16.5 based simulation tool and result show that the switching losses using zero-voltage-switching technique is less compared to conventional buck converter.
ABSTRAK

Projek ini mencadangkan rekabentuk litar pensuisan voltan-sifar penukar buck dc-dc untuk kegunaan penukar USB. Rekabentuk dan analisis dipersembahkan dengan litar bersepadu induktor. Litar penukar ini menukarkan voltan masukan 12V kepada voltan keluaran 5V dengan arus beban 1.5A pada frekuensi 50MHz. Litar dianalisa menggunakan perisian OrCAD Capture CIS dan berdasarkan hasil daripada simulasi litar menunjukkan kehilangan kuasa pensuisan menggunakan teknik pensuisan voltan-sifar lebih rendah berbanding dengan penukar buck konvensional.
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<td>DC</td>
<td>Direct Current</td>
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<td>AC</td>
<td>Alternating Current</td>
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<td>PWM</td>
<td>Pulse Width Modulation</td>
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<td>BJT</td>
<td>Bipolar Junctions Transistor</td>
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<td>IGBT</td>
<td>Insulated Gate Bipolar Junction</td>
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<td>MOSFET</td>
<td>Metal Oxide Semiconductor Field-Effect Transistor</td>
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CHAPTER I

INTRODUCTION

This chapter will focus on the brief of the project to be carried out. The important overview or description including the problem statement, project objectives, and project scopes are well emphasized in this part.

1.1 Overview

Recently, numerous soft switching techniques for the switching converters have been proposed in order to solve issues of conventional hard switching converters. These techniques can reduce the switching losses or in ideal case, it can achieve zero switching losses.

This report will investigate the ZVS operation with DC-DC resonant buck converter. A resonant conversion technology has been extensively used in power supplies in the consumer arena, with buck converters improving power supply efficiency. The principle of resonant conversion is to reduce the turn on losses of the power switch in a topology.

Advanced in resonant power conversion technologies propose alternative solutions to a conflicting set of square wave conversion design goals. An increasing challenge can be witnessed by emerging resonant technologies, primarily due to their lossless switching merits. The intent of this project is to unravel the details of zero voltage switching via comprehensive analysis of the timing intervals and relevant voltage and current waveforms.

The concept of resonant, ‘lossless’ switching is not new, most noticeably patented by one individual [1] and publicized by another at various power conferences [2,3]. Numerous efforts focusing on zero current switching ensued, first
perceived as the likely candidate for tomorrow’s generation of high frequency power converters [4,5,6,7,8]. In theory, the on off transitions occur at a time in the resonant cycle where the switch current is zero, facilitating zero current, hence zero power switching. And while true, two obvious concerns can impede the quest for high efficiency operation with high voltage inputs.

By nature of the resonant tank and zero current switching limitation, the peak switch current is significantly higher than its square wave counterpart. In fact, the peak of the full load switch current is a minimum of twice that of its square wave kin. In its off state, the switch returns to a blocking a high voltage every cycle. When activated by the next drive pulse, the MOSFET output capacitance is discharged by the FET, contributing a significant power loss at high frequencies and high voltages. Instead, both of these losses are avoided by implementing a zero voltage switching technique [9,10].

1.2 Problem statement

The importance of switching frequency quickly becomes apparent to systems designers bringing regulated power to on-board semiconductor devices. Accordingly, higher power density regulators with minimal PCB footprints have evolved; these use the latest in IC integration, MOSFETs and packaging. However, even these struggle keep pace with the continuing stream of newer, more powerful devices.

With these pressures, it is tempting to increase regulator frequency, as this reduces the size and board footprint requirements of the associated passive devices inductors, capacitors and resistors. Conventional thinking, based upon classic hard-switching PWM regulators, is that as frequency increases, then so do switching losses. This is because, in these topologies, regulator MOSFETs incur losses every time they switch, so a higher switching frequency leads directly to higher losses. These inefficiencies are primarily due to high side losses during turn on, Miller gate charge and body diode conduction losses. To make things worse, conventional topologies further magnify the losses as higher input voltages are converted or regulated.

These losses introduce a practical limit for the switching frequency of conventional converters and regulators. There is, however, a solution: devices that use the zero-voltage-switching (ZVS) topology do not suffer from losses in the same
way as conventional designs, allowing them to operate at higher frequencies, which in turn improves performance and dramatically reduces the size of external filter components. Converters and regulators using ZVS also reduce the additional losses associated with large step-down ratios in PWM topologies.

Unlike conventional regulators that rely on hard switching topology, ZVS uses soft switching; this accounts for its improved efficiency and higher density performance. Higher frequency operation not only reduces the size of passive components but also reduces the burden on external filtering components and allows for fast dynamic response to line and load transients.

1.3 Objective of the project

The main objective of this research is to design a zero-voltage-switching DC-DC buck converter and compare the results obtained to the hard-switched DC-DC buck converter. In theory, the resonant converter should yield a more compact design with the capability of power switching losses. A resonant converter will be measured and will be compared to the hard-switched converter to see if the common belief that soft-switched converter have reduce power switching losses. Based on result, a recommendation will be made as to whether the design is a suitable for a USB adapter application.

The objectives of this project were:

I. To derive zero-voltage-switching DC-DC converter using mathematical calculation
II. To perform the model of zero-voltage-switching DC-DC converter
III. To analyze the switching power losses of the resonant converter
IV. To compare the result of resonant converter with the hard-switching converter

1.4 Scope of the project

The scope of his project is to design the zero-voltage-switching DC-DC buck converter in order to reduce the switching losses that happen in conventional hard
switching. Scope is necessary to make sure that the objective of project will be achieved. The other scope and specification of this project is:

I. Resonant converter of this project has input voltage 12V and output voltage 5V
II. Model for this resonant converters will be develop using ORCAD Capture CIS simulation
III. Verification of this resonant converter includes switching power losses, input and output current and voltage waveform
CHAPTER II

RESONANT CONVERTER REVIEWS

This chapter will focus on the theoretical and history of converter. Literature review has been conducted prior to undertaking this project to obtain the information of the method that used by other researchers on the same topic.

2.1 Hard switching and soft switching background

Advances in power electronics in the last few decades have led to not just improvements in power devices, but also new concepts in converter topologies and control. In the 1970’s, conventional PWM power converters were operated in a switched mode operation. Power switches have to cut off the load current within the turn-on and turn-off times under the hard switching conditions. Hard switching refers to the stressful switching behavior of the power electronic devices. The switching trajectory of a hard-switched power device is shown in Figure 2.1. During the turn-on and turn-off processes, the power device has to withstand high voltage and current simultaneously, resulting in high switching losses and stress. Dissipative passive snubbers are usually added to the power circuits so that the dv/dt and di/dt of the power devices could be reduced, and the switching loss and stress be diverted to the passive snubber circuits. However, the switching loss is proportional to the switching frequency, thus limiting the maximum switching frequency of the power converters. Typical converter switching frequency was limited to a few tens of kilo-Hertz (typically 20kHz to 50kHz) in early 1980’s. The stray inductive and capacitive components in the power circuits and power devices still cause considerable transient effects, which in turn give rise to electromagnetic interference (EMI) problems.
Figure 2.1 - Typical Switching Trajectories of Power Switches.

Figure 2.2 shows ideal switching waveforms and typical practical waveforms of the switch voltage. The transient ringing effects are major causes of EMI.

Figure 2.2 - Typical Switching Waveforms

In the 1980’s, lots of research efforts were diverted towards the use of resonant converters. The concept was to incorporate resonant tanks in the converters to create oscillatory (usually sinusoidal) voltage and/or current waveforms so that zero voltage switching (ZVS) or zero current switching (ZCS) conditions can be
created for the power switches. The reduction of switching loss and the continual improvement of power switches allow the switching frequency of the resonant converters to reach hundreds of kilo-Hertz (typically 100kHz to 500kHz). Consequently, magnetic sizes can be reduced and the power density of the converters increased. Various forms of resonant converters have been proposed and developed. However, most of the resonant converters suffer several problems. When compared with the conventional PWM converters, the resonant current and voltage of resonant converters have high peak values, leading to higher conduction loss and higher V and I ratings requirements for the power devices. Also, many resonant converters require frequency modulation (FM) for output regulation. Variable switching frequency operation makes the filter design and control more complicated.

In late 1980’s and throughout 1990’s, further improvements have been made in converter technology. New generations of soft-switched converters that combine the advantages of conventional PWM converters and resonant converters have been developed. These soft-switched converters have switching waveforms similar to those of conventional PWM converters except that the rising and falling edges of the waveforms are ‘smoothed’ with no transient spikes. Unlike the resonant converters, new soft-switched converters usually utilize the resonance in a controlled manner. Resonance is allowed to occur just before and during the turn-on and turn-off processes so as to create ZVS and ZCS conditions. Other than that, they behave just like conventional PWM converters. With simple modifications, many customized control integrated control (IC) circuits designed for conventional converters can be employed for soft-switched converters. Because the switching loss and stress have been reduced, soft-switched converter can be operated at the very high frequency (typically 500kHz to a few Mega-Hertz). Soft-switching converters also provide an effective solution to suppress EMI and have been applied to DC-DC, AC-DC and DC-AC converters. This chapter covers the basic technology of resonant and soft-switching converters. Various forms of soft-switching techniques such as ZVS, ZCS, voltage clamping, zero transition methods etc. are addressed. The emphasis is placed on the basic operating principle and practicality of the converters without using much mathematical analysis.
2.2 Resonant converter overview

A soft-switching converter (also called a resonant converter) is broadly defined as a converter where additional switches, diodes, capacitors and/or inductors are added to create a zero voltage or current condition by LC resonance action before the switch is operated. All resonant DC-DC converters can be placed in either of the two following classes of converters:

I. Load-resonant converters, and
II. Resonant-switch converters

Load-resonant converters control the power flow to the load by resonant tank impedance, which is in turn controlled by the switching frequency or the resonant frequency [11]. Having to design a control system that changes the switching frequency as a function of loading is difficult and, as such, these classes of converters are not attractive for DC ZEDS. The component-fixed resonant frequency is even more difficult to change.

Conversely, resonant-switch converters use an LC resonance to shape the voltage and current waveforms to minimize the turn-on and turn-off losses. Reference [11] further divides the resonant-switch converter into the following categories:

I. Zero-Voltage Switching (ZVS)
II. Zero-Current Switching (ZCS)

Resonant-switch converters have the advantage that they are less sensitive to load changes and fixed-frequency Pulse-Width-Modulation (PWM) control schemes can be used on a number of these topologies.

2.2.1 Zero Current Switches

In a ZC resonant switch, an inductor \( L_r \) is connected in series with a power switch \( S \) in order to achieve zero-current switching (ZCS). If the switch \( S \) is a unidirectional switch, the switch current is allowed to resonate in the positive half cycle only. The resonant switch is said to operate in half-wave mode. If a diode is connected in anti-parallel with the unidirectional switch, the switch current can flow in both directions.
In this case, the resonant switch can operate in full-wave mode. At turn-on, the switch current will rise slowly from zero. It will then oscillate, because of the resonance between $L_r$ and $C_r$. Finally, the switch can be commutated at the next zero current duration. The objective of this type of switch is to shape the switch current waveform during conduction time in order to create a zero-current condition for the switch to turn off [12].

### 2.2.2 Zero Voltage Switches

In a ZV resonant switch, a capacitor $C_r$ is connected in parallel with the switch $S$ for achieving zero-voltage switching (ZVS). If the switch $S$ is a unidirectional switch, the voltage across the capacitor $C_r$ can oscillate freely in both positive and negative half-cycle. Thus, the resonant switch can operate in full-wave mode. If a diode is connected in anti-parallel with the unidirectional switch, the resonant capacitor voltage is clamped by the diode to zero during the negative half-cycle. The resonant switch will then operate in half-wave mode. The objective of a ZV switch is to use the resonant circuit to shape the switch voltage waveform during the off time in order to create a zero-voltage condition for the switch to turn on [12].

### 2.2.3 Comparisons between ZCS and ZVS

ZCS can eliminate the switching losses at turn off and reduce the switching losses at turn on. As a relatively large capacitor is connected across the output diode during resonance, the converter operation becomes insensitive to the diode’s junction capacitance. When power MOSFETs are zero-current switched on, the energy stored in the device’s capacitance will be dissipated. This capacitive turn-on loss is proportional to the switching frequency. During turn on, considerable rate of change of voltage can be coupled to the gate drive circuit through the Miller capacitor, thus increasing switching loss and noise. Another limitation is that the switches are under high current stress, resulting in higher conduction loss. However, it should be noted that ZCS is particularly effective in reducing switching loss for power devices (such as IGBT) with large tail current in the turn-off process. ZVS eliminates the capacitive turn-on loss. It is suitable for high-frequency operation. For single-ended configuration, the switches could suffer from excessive voltage stress, which is
proportional to the load. The maximum voltage across switches in half-bridge and full bridge configurations is clamped to the input voltage. For both ZCS and ZVS, output regulation of the resonant converters can be achieved by variable frequency control. ZCS operates with constant on-time control, while ZVS operates with constant off-time control. With a wide input and load range, both techniques have to operate with a wide switching frequency range, making it not easy to design resonant converters optimally.

2.3 Zero Voltage Switching research reviews

In year 1999, the authors in references [13], Ashton and J.G. Ciezki conducted a comprehensive study of many different resonant-switch DC-DC converters and ranked them on their ease of control, estimated efficiency and simplicity. Converter topologies that simply added external components to the classical converter illustrated in Figure 2.3 received consideration, as did converters requiring a constant or varying switching frequency control scheme.

![Figure 2.3- Classical Buck (DC-DC Converter)](image)

In year 1999, the author in reference [15], Turner C. described that the Joung converter has the advantage that no resonant components are placed in series with the main switch and that the topology is a simple extension of the basic buck converter. The converter also has the added benefit that constant frequency PWM control is feasible. The circuit diagram of the Joung converter circuit as shown in Figure 2.4.
In year 2001, the author in reference [16], Bryan D. Whitcomb was study on the design and implementation of a high-power resonant dc-dc converter module for a reduce-scale prototype integrated power system in order to verify improvements in efficiency and Electromagnetic Interference (EMI). The converter was designed using the concept of Joung converter circuit with the IGBT switch so that it met the specifications set forth by ESAC for the reduced-scale IPS laboratory. His research focused on the Ships Service Converter Module (SSCM), which in its most basic form, is simply a buck chopper DC-DC converter.

Author in reference [17], Per Karlsson, Martin Bojrup, Mats Alakula and Lars Gertmar was study on zero voltage switching converter. The investigation discusses the concept of resonant and quasi-resonant DC link converter. The application of their research is implemented and tested in a battery charger. For the resonant DC link converter, one resonant circuit is used to provide soft switching for the entire converter. As the name resonant DC link indicates the DC link is forced to oscillate. The basic three phase resonant DC link converter is shown in Figure 2.5.

Figure 2.4- Joung Resonant Converter Topology
In quasi-resonant DC link converters the link voltage is not continuously oscillating, but can be triggered to perform a resonant cycle. The circuit investigated here, is the passively clamped two switch quasi-resonant converter. Figure 3.6 show, the passively clamped two switch quasi-resonant DC link is used in a battery charger application. The inductor L2 and the transistors S1 and S2 together with the diodes D1 and D2 form the trig on demand ability. The transformer L1/L3 and the diode D3 are part of a voltage clamping network used to limit the resonant link voltage.

At the end of their research, they found that the measurements show that the converter losses increase compared to the hard-switched counterpart at a switching frequency of 5 kHz. Despite the low switching frequency, the measurements show that the commanded switching are delayed, causing low order harmonics.

In year 2009, the author in reference [18], Rahul Shrivastava and Gupta Saurabh was come out with research title Study and Design of a Zero Voltage Switched Boost Converter. They were study the theoretical circuit operation of zero voltage switching over the basic premise of boost converters (step-up dc chopper circuits). The paper presents the concept of Zero Current Switching Technique and
Zero Voltage Switching Technique. ZVS Boost converter provides good zero voltage switching conditions for both the transistor and the diode. ZVS boost converter generates dc voltage which can be applied in power supply systems where high energy efficiency is required.

Figure 2.7- ZVS Boost Converter circuit diagram

In year 2013, the author in reference [20], MD. Imran and K. Chandra Mouli, had proposed the Simulation of a Zero-Voltage-Switching and Zero-Current-Switching Interleaved Boost and Buck Converter. By using the interleaved approach, the topology not only decreases the current stress of the main circuit device but also reduces the ripple of the input current and output voltage. Besides that, the soft-switching interleaved converter also can reduce the size and cost in the common soft-switching module. The author using MATLAB/Simulink simulation to obtain the simulation result in their research.
CHAPTER III

METHODOLOGY

This chapter will describe the method that will be used for this project in order to achieve the desire project. This project development is divided into two parts. The first part represented the proposed resonant converter design. The design of the converter has to be adjusted to obtain the best performance of the system. The specifications for resonant converter were listed and resonant component values were calculated based on the specifications. The results then have to be analyzed to ensure the best switching power losses of converter. If the results are not satisfied, the controllers should be adjusted due to their parameters. The next step is followed by selection of the semiconductor switch. The concept of a semiconductor switch was presented and different families will be study.

3.1 Proposed converter design

As stated in Chapter 1, the goals of this thesis are to design a resonant soft-switching DC-DC converter. The resonant converter described in this thesis was designed, built and tested along with the hard-switched converter so that some meaningful comparisons could be documented. The specifications for the resonant converter are listed in Table (3-1).
Table 3-1: Resonant Converter Specifications

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Specifications</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input voltage ($V_i$)</td>
<td>12</td>
</tr>
<tr>
<td>Output voltage ($V_o$)</td>
<td>5</td>
</tr>
<tr>
<td>Switching frequency ($f_s$)</td>
<td>50MHz</td>
</tr>
<tr>
<td>Resonant capacitance ($C_r$)</td>
<td>0.315nF</td>
</tr>
<tr>
<td>Resonant inductance ($L_r$)</td>
<td>35nH</td>
</tr>
<tr>
<td>Output ripple capacitor ($C_o$)</td>
<td>31.5nF</td>
</tr>
<tr>
<td>Output ripple inductor ($L_o$)</td>
<td>3500nH</td>
</tr>
<tr>
<td>Load ($R_o$)</td>
<td>3.33</td>
</tr>
</tbody>
</table>

3.2 Resonant converter design

For the resonant buck converter, the designer must choose the correct component values so that proper circuit operation occurs. The components can be divided into two groups: the ones that make up the core of the soft-switching circuitry and the other components necessary for the classical hard-switched buck converter. The resonant converter is shown in Figure 3.1. The additional resonant components are contained inside the dashed area of Figure 3.1, where $C_r$ is simply the body capacitance of the main switch.

![Figure 3-1: The proposed resonant converter circuit](image)

Since this converter design is much the same as the basic buck converter with an additional diode, resonant capacitors and a resonant inductor, the proper place to
initiate the design process is the selection of the proper values of the main inductor and capacitor.

### 3.2.1 Selection of the resonant inductor and the resonant capacitor

The determination of the resonant inductor and resonant capacitor is well understood and the requirements are based on switching frequency, load resistance, and output current. The load resistance is defined by dividing the desired output voltage by the output current

$$R = \frac{V_o}{V_i} = \frac{5}{1.5} = 3.333\Omega$$  \hspace{1cm} (3.1)

The $r$ and $fs/fo$ is defined by relationship between $Vo/Vin$ curve in Figure 3.2

From the graph, when $Vo/Vin = 0.416$, $r=0.1$ and $fs/fo=0.33$.

Then, the load impedance is defined by dividing the load resistance by the $r$

$$Z_o = \frac{R}{r} = \frac{3.333}{0.1} = 33.33\Omega$$  \hspace{1cm} (3.2)
Frequency output, $f_o$ is defined by,

$$f_o = f_{\text{switching}} = \frac{50M}{0.33} = 151.515\text{MHz}$$

(3.3)

Then,

$$\omega_o = 2\pi f_o = 2\pi \times 151.515\text{MHz} = 951.998\text{M}$$

(3.4)

The resonant inductance is found to be

$$L_r = \frac{Z_o}{\omega_o} = \frac{33.33}{951.998\text{M}} = 34.979\text{nH}$$

(3.5)

Then, the resonant capacitance is found by using equation (3.6)

$$Z_o = \sqrt{\frac{L_r}{C_r}}$$

(3.6)

$$C_r = \frac{L_r}{Z_o^2} = \frac{34.979\text{nH}}{33.33^2} = 0.0315\text{nF}$$

3.2.2 Mode of operation

During the steady-state operation and if the bulk energy storage components, $L_f$ and $C_f$, are large, the analysis can be simplified by assuming the current through the bulk inductor is a current sink. The circuit can be redrawn, simplified and can be considered in four equivalent circuits that depend on the switching device SW and diode's on- and off-states as shown in Figure 3.3. There are four states of operation and the time domain waveform is shown in Figure 3.3
a) Capacitor charging stage - switch SW is turned off at $t_0$. Input current $I_{Lr}$ rises linearly and is governed by

$$v_{Cr} = \frac{I_o(t-t_0)}{C_r}$$  \hspace{1cm} (3.7)

When $V_{Cr}$ increases to $V_{in}$, the voltage across $DF$ becomes positive and it is in forward bias.

The duration of $t_1$ at this stage is,

$$T_{d1} = C_r \frac{V_{in}}{I_o}$$  \hspace{1cm} (3.8)

And the boundary condition is,

$$v_{C_r} = V_{in}$$  \hspace{1cm} (3.9)

b) Resonant state: $Lr$ and $Cr$ resonate and $DF$ is on,

$$v_{Cr} = V_{in} + ZI_o \sin \omega_o(t-t_1)$$  \hspace{1cm} (3.10)

$$i_{Lr} = I_o \cos \omega_o(t-t_1)$$  \hspace{1cm} (3.11)

The switch SW consists of transistor $T$ and diode $D$. When resonant voltage $V_{cr}$ reaches zero, it cannot be reversed because the anti-parallel diode $D$ of the switch conducts. The transistor $T$ of the switch SW can be turned on after that to achieve zero-voltage switching. The boundary condition is,
\[ v_G(t) = 0 \]  \hspace{1cm} (3.12)

At \( t_2 \),
\[ i_{Lr} = I_o \cos \alpha \]  \hspace{1cm} (3.13)

c) Inductor recovering stage: (Fig. 4c) Resonance stops, \( L_r \) begins to be charged by the input voltage \( V_{in} \)
\[ i_{Lr} = I_o \cos \alpha + \frac{V_{in}(t - t_2)}{L_r} \]  \hspace{1cm} (3.14)

This state finishes when \( i_{Lr} \) reaches the value of output current \( I_o \), DF no longer conducts because its current is now all conducted by \( L_r \)
The duration is,
\[ T_{d2} = L_r \frac{I_o \cos \alpha}{V_{in}} \]  \hspace{1cm} (3.15)

And boundary condition,
\[ i_{Lr} = I_o \]  \hspace{1cm} (3.16)

d) Free-wheeling stage- Output current freewheels through \( L_r \) and switch SW. This stage finishes when the transistor turns off again at \( t_4 \). \( t_4 \) is the same as \( t_0 \) in next cycle.
\[ T_{d4} = T_s - T_{d1} - T_{d2} - T_{d3} \]  \hspace{1cm} (3.17)

where \( T_s \) is the period of the switching cycle. Mathematical calculation for \( t_1 \), \( t_2 \), \( t_3 \), and \( T_s \) shown in Equation 3.18 to 3.22.

\[ t_1 = \frac{V_s C_r}{I_o} = \frac{12(0.0315\pi)}{1.5} = 0.252\text{ns} \]  \hspace{1cm} (3.18)

\[ t_2 = \frac{1}{\omega_o} \left[ \sin^{-1} \left( \frac{V_s}{I_o Z_o} \right) + \pi \right] + t_1 \]  \hspace{1cm} (3.19)

\[ t_2 = \frac{1}{951.998M} \left[ \sin^{-1} \left( \frac{12}{1.5 	imes 33.33} \right) + \pi \right] + 0.252\text{ns} \]
\[ t_2 = 3.414\text{ns} \]

\[ t_3 = \left[ \frac{L_r I_o}{V_s} \right] \left[ 1 - \cos (\omega_o (t_2 - t_1)) \right] + t_2 \]  \hspace{1cm} (3.20)

\[ t_3 = \left[ \frac{34.979\text{nx} \times 1.5}{12} \right] \left[ 1 - \cos (951.998M(3.414\text{ns} - 0.252)) \right] + 3.414\text{ns} \]
t_3 = 12.351\text{ns}

\[ f_{\text{max}} = \frac{1}{t_3} = \frac{1}{12.351\text{ns}} = 80.966\text{MHz} \] \hspace{1cm} (3.21)

\[ T = \frac{1}{f_s} = \frac{1}{50\text{M}} = 20\text{ns} \] \hspace{1cm} (3.22)

### 3.3 Selection of the semiconductor switch

Each of these devices will now be assessed and the advantages and disadvantages will be analyzed to determine the best device for the resonant converter in this thesis. Figure 3.4 illustrates the more common devices in use today and their relative voltage, current and switching frequency capabilities.

![Figure 3.4 - Summary of Semiconductor Switching Devices (from ref. [11])](image)

So far in all of the analysis, it was assumed that both the main and auxiliary switches were ideal devices with the following characteristics:

I. Zero current flow when blocking large voltages,

II. Zero voltage drop when conducting large currents,

III. Device can be switched "on" and "off" instantaneously when triggered,

IV. Triggering device requires no power.
As expected, no devices exist with these characteristics, but different device families are available with relative advantages over each other in these areas. Currently, the controllable turn on/off devices in use today for power converters are the Metal-Oxide Field Effect Transistor (MOSFET), Bipolar Junction Transistor (BJT) and the Insulated Gate Bipolar Transistor (IGBT). Many other families based on the thyristor exist, but they were not considered, since they are typically a controlled-turn-on device only. Other variants of thyristors such as the Gate-Turn-Off (GTO) are controlled-turn-off devices, but were not considered due to their slow switching speeds.

All real switching devices will have switching characteristics similar to Figure 3.5. During one complete cycle of the switch there will be energy loss during turn-on, turn-off and steady-state conduction. When a gating signal is applied, a real semiconductor switch will not react instantaneously and will have delays associated with charge build-up in the device.

Figure 3.5- Generalized Switching Waveform
The reverse recovery of a diode circuit will also increase the turn-on losses, because the switch current must reach the diode current with system voltage still applied across the switch before the diode is starved of current. The main switch, $M_1$, paired with diode $D_m$ in Figure 3.6 illustrates this turn-on loss. When the gate signal is applied, $D_m$ is forward biased and $M_1$ has $V_{dc}$ across its terminals. The current rapidly builds up through $M_1$, but the voltage across $M_1$ remains at $V_{dc}$ until $D_m$ is starved of current and becomes reversed biased.

Following turn-on, the switch undergoes a period of steady-state conduction loss due to its internal resistance and current flow. Typical power transistors will have voltage drops of 2 to 5 volts during a full-on conduction stage and this will add to the total power loss.

![Figure 3.6- Zero-voltage-switching converter schematic](image)

Finally, when a switch is gated "off," charge must be removed from the gate/base region and in some cases recombination has to occur before the current decays to zero. As in the turn-on case, full voltage will be across the device while this transition occurs, thus power loss will again happen.

The switching losses can be estimated by treating the total area under the voltage-current curve of Figure 3.5 as a triangle. The following energy loss equation can be written based on this simple geometry.
\[ W_{on} = \frac{1}{2} V_d I_o t_{c(on)} \] \hspace{1cm} (3.23)

\[ W_{off} = \frac{1}{2} V_d I_o t_{c(off)} \] \hspace{1cm} (3.24)

\[ W_{conduction} = V_d I_o t_{on} \] \hspace{1cm} (3.25)

With a switching frequency of \( f_s \), the above amounts of energy loss occur \( f_s \) times in a second and thus:

\[ P_{on} = \frac{1}{2} V_d I_o t_{c(on)} f_s \] \hspace{1cm} (3.26)

\[ P_{off} = \frac{1}{2} V_d I_o t_{c(off)} f_s \] \hspace{1cm} (3.27)

\[ P_{conduction} = V_d I_o t_{on} f_s \] \hspace{1cm} (3.28)

and

\[ P_{total} = P_{on} + P_{off} + P_{conduction} \] \hspace{1cm} (3.29)

### 3.3.1 Bipolar Junction Transistor (BJT)

The first device considered was the Bipolar Junction Transistor (BJT). The BJT is a current-controlled device with a negative temperature coefficient. A negative temperature coefficient is considered a disadvantage as the device is prone to thermal runaway problems and makes device paralleling difficult. The base region in power transistors tends to be wider, which drastically reduces the current gain \( \beta \) in comparison to a small signal transistor [19]. The result is that the base current, \( i_b \), is no longer insignificant and is a large fraction of the emitter current. The large base current makes the design of the driver circuitry difficult. To increase \( \beta \), power transistors are connected in a Darlington configuration or even a tripleton configuration, at the cost of decreased switching speeds [11]. The development of power BJTs has all but stopped due to the advancement of voltage-controlled transistors such as the IGBT.
3.3.2 Insulated Gated Bipolar Transistor (IGBT)

The next device considered for this design is IGBT. The IGBT switch is a bipolar transistor with a MOSFET transistor driving the base in a Darlington configuration. The IGBT is a voltage-controlled device with a main body region similar to a BJT transistor. During the blocking state, the gate voltage is below the threshold voltage and the circuit voltage is dropped across the depletion region with little leakage current flowing. When the gate voltage exceeds the threshold, an inversion layer forms beneath the gate and shorts the n⁺ region to the n⁻ region. The current that flows causes hole injection from the p⁺ region into the n⁻ region. The injected holes continue to the p-body region via drift and diffusion, and the excess holes attract electrons from the source connection. This process continues until the gate voltage returns to a value less than the threshold voltage, and the device transitions to the off-state.

3.3.3 MOSFET

The MOSFET is a three-terminal (gate, drain, and source) fully-controlled switch. The gate/control signal occurs between the gate and source, and its switch terminals are the drain and source. The gate itself is made of metal, separated from the source and drain using a metal oxide. This allows for less power consumption, and makes the transistor a great choice for use as an electronic switch or common-source amplifier. A physical makeup of MOSFET is illustrated in Figure 3.7.
REFERENCES


17. Department of Industrial electrical Engineering and Automation, Lund University “Zero Voltage Switching Converters” by Per Karlsson, Martin Bojrup, Mats Alaküla and Lars Gertmar,
