POLITECNICO
MILANO 1863

SCHOOL OF INDUSTRIAL AND INFORMATION ENGINEERING

MASTER OF SCIENCE IN ELECTRICAL ENGINEERING

Resonant converters with insulating transformers and wide voltage variation

Supervisor: Prof. Morris Brenna

Author: Abdelrahman Ihabeldin Ali Ahmed
Matricola: 10635846 - (904163)

Academic Year 2019 - 2020

Contents

List of figures

Abstract

With great advances of power semiconductor switching devices such as MOSFETs, IGBTs and ESBTs as well as high-frequency passive circuit components, the leading development of the high frequency resonant pulse inverter type switching mode DC-DC power conversion circuits and systems have attracted special interest for high voltage DC power applications

This paper presents a study on Resonant converters with insulating transformers and with wide voltage variation. It will be composed of several chapters explaining the methodology of operation of these converters and the insulating transformers. In one chapter, the history of Resonant converters will be discussed and why they are selected over the regular switching dc-dc converter, the switching modes of the converter (ZVS & ZVC). The following chapter will be a discussion about the high frequency transformers, their concept and topology. Then I will speak about Galvanic Insulations and its types along with its Advantages. Lastly an Application on charging infrastructures in Electrical Vehicles.

Astratto

Con i grandi progressi dei dispositivi di commutazione a semiconduttore di potenza come MOSFET, IGBT ed ESBT nonché componenti di circuiti passivi ad alta frequenza, lo sviluppo principale dei circuiti e dei sistemi di conversione di potenza CC-CC con modalità di commutazione ad impulsi risonanti ad alta frequenza ha suscitato particolare interesse per applicazioni di alimentazione CC ad alta tensione.Questo articolo presenta uno studio sui convertitori risonanti con trasformatori isolanti e con ampia variazione di tensione. Sarà composto da diversi capitoli che spiegano la metodologia di funzionamento di questi convertitori e dei trasformatori isolanti. In un capitolo, verrà discussa la storia dei convertitori risonanti e il motivo per cui sono stati selezionati rispetto al normale convertitore cc-cc a commutazione, le modalità di commutazione del convertitore (ZVS e ZVC). Il capitolo seguente sarà una discussione sui trasformatori ad alta frequenza, il loro concetto e la topologia. Poi parlerò degli isolamenti galvanici e dei suoi tipi insieme ai suoi vantaggi. Infine un'applicazione sulle infrastrutture di ricarica dei veicoli elettrici.

Acknowledgement

First of all, all praise and gratitude to Almighty Almighty Allah, who enabled me to complete this thesis. I would not have been able to finish without his blessings.

I am very thankful to Politecnico di Milano, for allowing me this opportunity, and Professor Morris Brenna whose encouragement, guidance, and support enabled me to develop the thesis project.

Special thanks to my Father PhD Ihabeldin Ali for his guidance and support during the thesis work and to my friend Eng. Kareem Hesham for his help.

Most importantly, I would like to express my eternal gratitude to my family, my mother, brother and sister, my fiancée, and my friends for their continuous support during this journey.

Finally, I would like to extend my gratitude to whoever provided me with any kind of help no matter how big or small.

1.Introduction

1.1 Historical Background

With great advances of power semiconductor switching devices such as MOSFETs, IGBTs and ESBTs as well as high-frequency passive circuit components, the leading development of the high frequency resonant pulse inverter type switching mode DC-DC power conversion circuits and systems have brought special interests for high voltage DC power applications

The “switching” dc – dc converter suffers from high switching loss and decreasing reliability. Even increasing power densities has been limited by the size of both reactive elements and the isolation transformer. Although component sizes tend to decrease, system switching losses are proportional to the frequency obtained in a given circuit with an increase in the switching frequency. The high frequencies are a key to realizing multiple benefits of high-power density and good transient response.

The use of soft-switching techniques mitigates switching loss problems and allow a major increase in the converter’s switching frequency. Most of the switching losses found in pulse width modulation (PWM) converters are removed by resonant converters (RCs). Two methods are used to adjust the active device: zero current switching (ZCS) or zero voltage switching (ZVS).The main criteria included in the specification of any dc/dc converter are the line regulation, load regulation, stability and response time.

The supply voltage and the load regulation have a large range of variability, so the controller would be designed to provide suitable behavior in the converter’s many working conditions. The design of a controller for the converter in the resonant converter is usually done using classical control methods. If there can be a significant change in the operating point, changes in the linear model must also be considered. One way to fulfill all the specifications is to study the design of the controller for worst conditions. Rectifiers normally used diode or thyristor line commutated circuits to produce uncontrolled or controlled DC outputs. However these types of circuits have low or weak waveforms and power factor input current, and as power quality requirements such as IEEE Standard 519 or the aircraft current harmonic limits standard have been tightened, the use of these circuits in many applications has become unpreferable and undesirable.

Within the different types of resonant converters, the LC series resonant converter is the simplest and most common resonant converter. Where the rectifier's load network is put in series with the resonant network LC. Although much work has been done since its introduction in the 1990s on the LLC resonant converter topology. Compared to the conventional LC series resonant converters, LLC resonant converters demonstrate many advantages, such as narrow frequency variation over a wide range of load and input variations and zero voltage switching even under no load conditions. For all power levels today, switched-mode power supplies based on resonant operating converter topologies are becoming increasingly interesting. Resonant topologies are typically applied when low EMI is needed or when the switching losses have to be reduced in order to allow higher frequencies for miniaturization.

In addition, resonant operation enables high frequency power transfer via a transformer. The LLC Series-resonant half-bridge converter is popular due to being a high efficiency dc-dc converter. In the constant output voltage mode, a half bridge parallel resonant converter running above resonance was analyzed by M. (1991) by Emsermann. Darryl J. Tschirhart and Praveen K. Jain (2010) offered secondary-side control of a constant frequency series resonant converter using dual-edge PWM. The Three-port Bi-directional DC-DC Converter Constant Switching Frequency Series Resonant was supplied by H. N. and Krishnaswami. (2008), Mohan. Henry W. Koertzent et al. (1995) provided the design of the Half-Bridge Series Resonant Converter for Induction Cooking.

1.2 Thesis outline:

- Chapter 1: Introduction
- Chapter 2: Resonant converters and its Applications
- Chapter 3: High Frequency Transformers
- Chapter 4: Resonant converters with Galvanic insulations
- Chapter 5: Application in charging infrastructures for EVs
- Chapter 6: Conclusion

2. Applications on Resonant Converters:

2.1. Resonant circuits

In power supply circuits, resonant circuits are used for many purposes, such as: high frequency voltage or current output, reduction of switching losses and switching noise.
A resonant circuit is a type of electric circuit that in response to externally applied energy, produces vibrations. Various electric circuits make use of resonant vibrations. Basically, a coil (L) and a capacitor (C) form a resonant circuit. There are two types of resonant circuits: resonant circuits in parallel, and resonant circuits in order. Parallel resonant circuits have efficient infinite impedance at the frequency of the resonant level, whereas series resonant circuits have zero impedance. "A parameter called quality factor" (Q factor) shows the energy loss of a resonant circuit. The higher the Q factor, the lower the energy loss rate and thus a more preferable resonator. Resonant circuits with a high Q factor might not go into resonance or, depending on their implementations, take a long period of time to become stable if their frequency deviates from the resonant frequency. In the case of a resonant inverter, the Q factor of the resonant circuit varies significantly with the load impedance. The The Q factor is expressed as the ratio of the load impedance to the LC impedance of such a resonant inverter, and is thus influenced by the load impedance. The resonant inverter's optimal topology depends on the impedance of the load. Three typical types of resonant circuits used in resonant inverters are shown in the figure below.

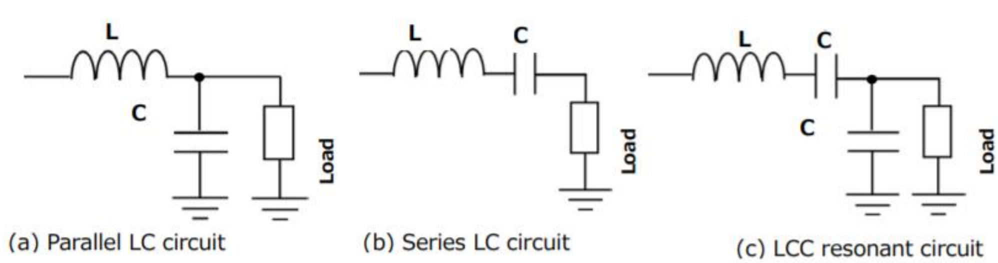


Figure 2.1 Resonant circuits

In Figure Figure 2.1(a) The parallel LC circuit of is used when a load has large impedance. This LC circuit goes into resonance properly if the impedance is very high, as it can be seen as similar to a circuit consisting only of an inductor L) and a capacitor (C). On the other hand, assuming that the load is minimal, C has little effect. As a result, this circuit does not go into resonance since it is only similar to a GND-linked inductor. The series LC resonant circuit with a lower impedance value in Figure 2.1(b) enables it to work in a manner similar to a resonant circuit consisting only of an inductor and a capacitor, thus providing a high Q. Conversely, Q becomes smaller if the impedance is high, rendering the series LC circuit unsuitable for series resonance. Figure 2.1(c) The LCC resonant circuit, another type of parallel and series LC circuits, functions as a resonant circuit, taking the impedance value into account. The optimum resonant circuit type must be chosen according to the load, as stated above.

2.2. Soft switching (reduction of switching losses and switching noise):

There are several techniques and methods for soft switching. Let us take for example the Partial resonance. Using the resonance phenomenon, it switches on and off power switching devices at zero voltage or zero current. The switching losses become considerably small in partial resonance because either voltage or current is zero as they turn on or off.

Zero-voltage crossing switching is known as zero-voltage switching (ZVS), while zero-current crossing switching is known as zero-current switching (ZCS).

Soft switching helps minimize switching losses caused by the switching of power devices. Furthermore, di/dt and dv/dt lamps are reduced and this is efficient in minimizing harmonic component switching and transient noise.

2.2.1. Comparisons Between ZCS and ZVS:

ZCS can neutralize turnoff switching losses and decrease turn-on switching losses. As a large capacitor is attached during resonance in the output diode, the converter operation becomes ineffective with the junction capacitance of the diode. As zero-current power MOSFETs are turned on the energy contained in the capacitance of the system will dissipate. This loss of capacitive turn-on is directly proportional to the frequency of the switching. Using the Miller capacitor, a significant rate of voltage shift may be coupled to the gate drive circuit during turn-on, which implies an increase in switching loss and noise. Another drawback is that the high stress to which the switches are exposed can cause greater loss of conduction. However, in the turn-off process, such as IGBT, it should be noted that ZCS is effective in reducing switching losses for power devices with a large tail current. ZVS, on the other hand, prevents the lack of capacitive turn-on. For high-frequency operations, it is acceptable. The switches could suffer from very high voltage stress in a single-ended configuration, which is proportional to the load. The maximum voltage across switches is shown to be secured to the input voltage in half-bridge and full-bridge configurations. The output regulation of the resonant converters can be achieved through variable frequency control for both ZVS and ZCS. With constant on-time control, ZCS operates, while ZVS operates with constant off-time control. Both techniques have to operate with a wide switching frequency range, with a wide input and load range, making it not easy to optimally design resonant converters.

2.2.2. Applications of soft switching

In addition to the turn-off loss of switching devices, soft switching helps minimize transformer leakage inductance loss and diode recovery loss. Soft switching helps reduce transformer leakage inductance loss and diode recovery loss, in addition to the turn-off loss of switching devices. It is important to consider both the benefits and disadvantages of soft switching to decide if it is an appropriate technique for your application. Although the advantages of soft switching are reduced switching losses and high-frequency noise, it has disadvantages such as an increase in the number of parts and a need for more complicated control. For switched-mode power supplies which have a high-frequency transformer, such as isolated DC-DC converters (LCC resonant converters), soft switching is used. As high-frequency transformers have leakage inductance, when the energy contained in the leakage inductance is dissipated by transistor switching, the performance of switched-mode power supplies is decreased. Soft switching also helps reduce the losses induced by the leakage and excitation inductances of a high-frequency transformer. Furthermore, soft switching results in a minimal increase in the number and cost of parts, as the leakage inductance of a high-frequency transformer can be used as a resonance inductor.

2.2.3. Soft-switching circuit topology

In the following section the basic circuit topology and the operation of partial resonance will be explained.

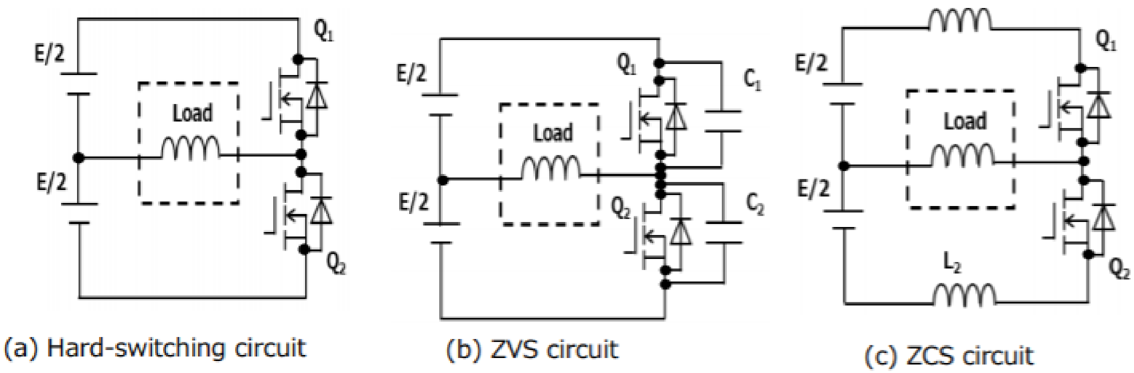


Figure 2.2 Simple half-bridge inverters.

The hard-switching circuit using typical switching techniques is shown in Figure (a). As shown in Figure (b), the ZVS topology connects C1 and C2 in parallel with switching devices. At zero voltage, this circuit achieves switching, resonating with load and other inductances. As shown in Figure (c), the ZCS topology links L1 and L2 in series with the switching devices. This circuit, resonating with load and other capacitances, achieves zero-current switching. Due to the reduced di/dt and dv/dt ramps during switching devices, the ZVS and ZCS methods help reduce switching losses, as well as the impact on the circuit and the noise generated by switching devices.

<u>ZVS operation</u> Turn-off operation

In Figure 2.2, Q1 and Q2 are MOSFETs, DQ1 and DQ2 are MOSFET body diodes, and C1 and C2 are the sums of the parasitic and external capacitances of Q1 and Q2, respectively. When Q2 turns off, the Q2 voltage begins to rise, causing the current flowing through Q2 to circulate to C2. Consequently, the C2 voltage (vQ) increases gradually. The voltage slope depends on the capacitance of C2. A higher C2 value results in a shallower vQ slope and therefore lower switching loss. Figure 2.3 shows the current and voltage waveforms for different C2 values.

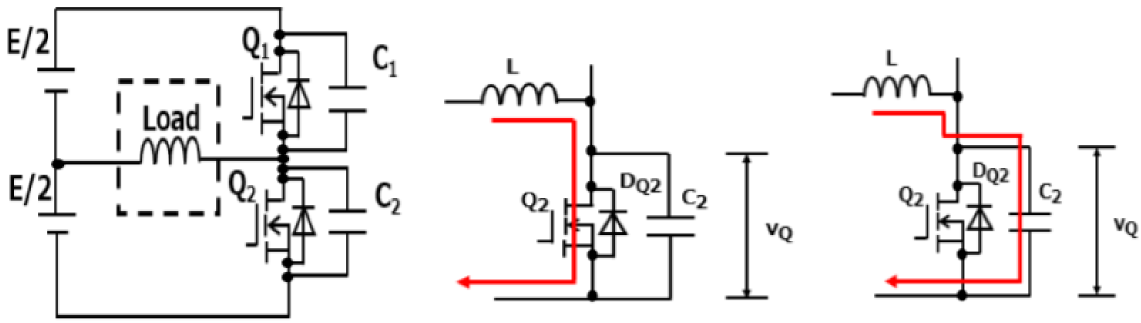


Figure 2.3 Turn-off operation

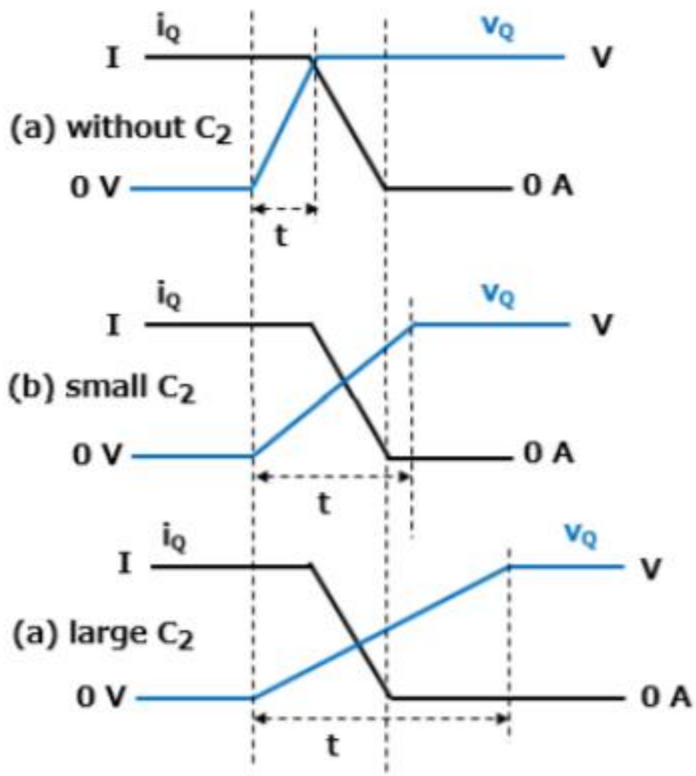


Figure 2.4 Turn off waveforms

Turn-on operation

Next, the following paragraphs discuss the switching device's ZVS turn-on. When Q1 is on and Q2 is off, Q2 turns on immediately after Q1 turns off, as described below. Immediately after Q1 switches off, the charge remains in capacitor C2. If it were activated in this state, Q2 would perform hard switching. It is necessary to remove the charge from C2 to allow diode DQ2 to conduct (Vq = Vf) to achieve soft switching and then turn on Q2. This sequence is demonstrated in Figure 2.4.

(a) Charge is removed from C2. To remove the charge from C2, the energy stored in L is used. (When Q1, a high side switching device, switches from on to off the current flows due to the energy stored in L as shown by the arrow in a].)
(b) When the charge is removed from C2 completely, the current passes through DQ2.
(c) While conducting DQ2, Q2 is turned on. Since VQ at this time is almost 0 V, zero-voltage switching (ZVS) occurs.
(d) <u>On state</u>: When the energy stored in load L is released, the current passing through Q2 reverses its direction, putting Q2 back in the normal "on" state, with current flowing from drain to source. At this time, Q2 is on, and its voltage (vQ) is nearly 0 V. Therefore, Q2 achieves soft switching.

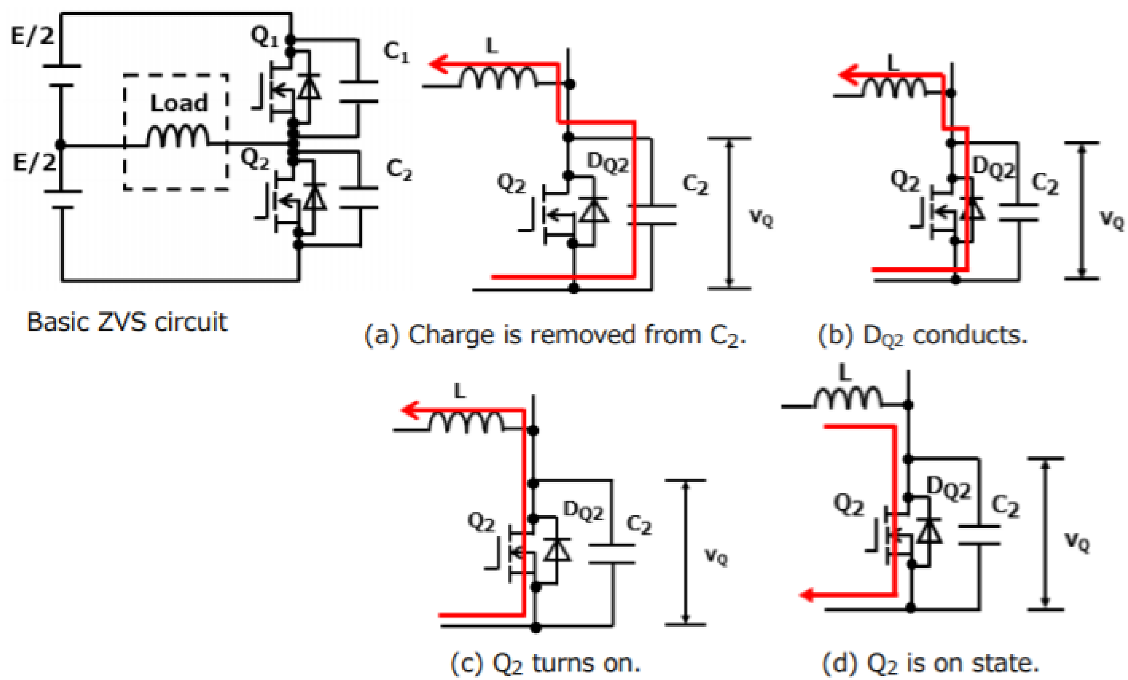


Figure 2.5 ZVS turn-on operation

2.3 LLC resonant converter

A DC-DC converter (power supply) requires downsizing, high efficiency, high output power and low EMI noise. To create such a power supply, a circuit topology that combines high-frequency switching and high efficiency is necessary. Previously, PWM (Pulse Width Modulation) control was commonly used for converters. However, the increase in switching frequency has caused problems such as increased switching loss due to turn-on and turn-off and generation of high-frequency noise.

A DC-DC converter (power supply) requires downsizing, high efficiency, high output power and low EMI noise. To create such a power supply, a circuit topology that combines high frequency switching and high efficiency is necessary. Previously, PWM (Pulse Width Modulation) control was commonly used for converters. However, the switching frequency increasing has caused problems such as increased switching loss due to turn-on and turn-off and generation of high-frequency noise.

An LLC resonant converter provides several desired characteristics such as low parts count, high efficiency and low noise, which combines current-resonant operation and soft switching (partial resonance). The name “LLC” is taken from the leakage inductance (Lr) and excitation inductance (Lm) of a transformer and capacitor (C) that are used to achieve resonance .Figure 2.6 shows an LLC resonant converter, that is composed of a half-bridge (i.e., a square-wave generator) consisting of two series MOSFETs (Q1 and Q2), a resonant capacitor (Cr), a transformer (T), two output rectifier diodes (D1 and D2), and an output capacitor (Co). In Figure 2.7, N1 represents the number of turns in the primary winding whereas N2 and N3, which are equal, represent the numbers of turns in the secondary windings. Lr is the leakage inductance of the transformer's primary winding. Generally, in an LLC resonant converter, a transformer with a low coupling coefficient is used to provide high leakage inductance such that the leakage inductance behaves as resonant inductance. In some cases, a separate inductor is connected in series with a transformer. Lm represents excitation inductance.

Even in the presence of input voltage variations, an LLC resonant converter makes it possible to reduce the control frequency range through the use of two different resonant frequencies: a fixed resonant frequency (fr) of an inductor-capacitor (Lr-Cr) pair and a resonant frequency (fm) derived from the inductor-capacitor pair ((Lr + Lm)-Cr) that varies with load Ro. The LLC resonant converter generates a square-wave voltage at a duty cycle close to 50%, which is converted into a nearly sinusoidal current with an LLC resonant circuit. Since the LLC resonant converter suppresses harmonics and allows ZVS soft-switching operation, it is widely used for applications requiring high efficiency and low EMI noise.

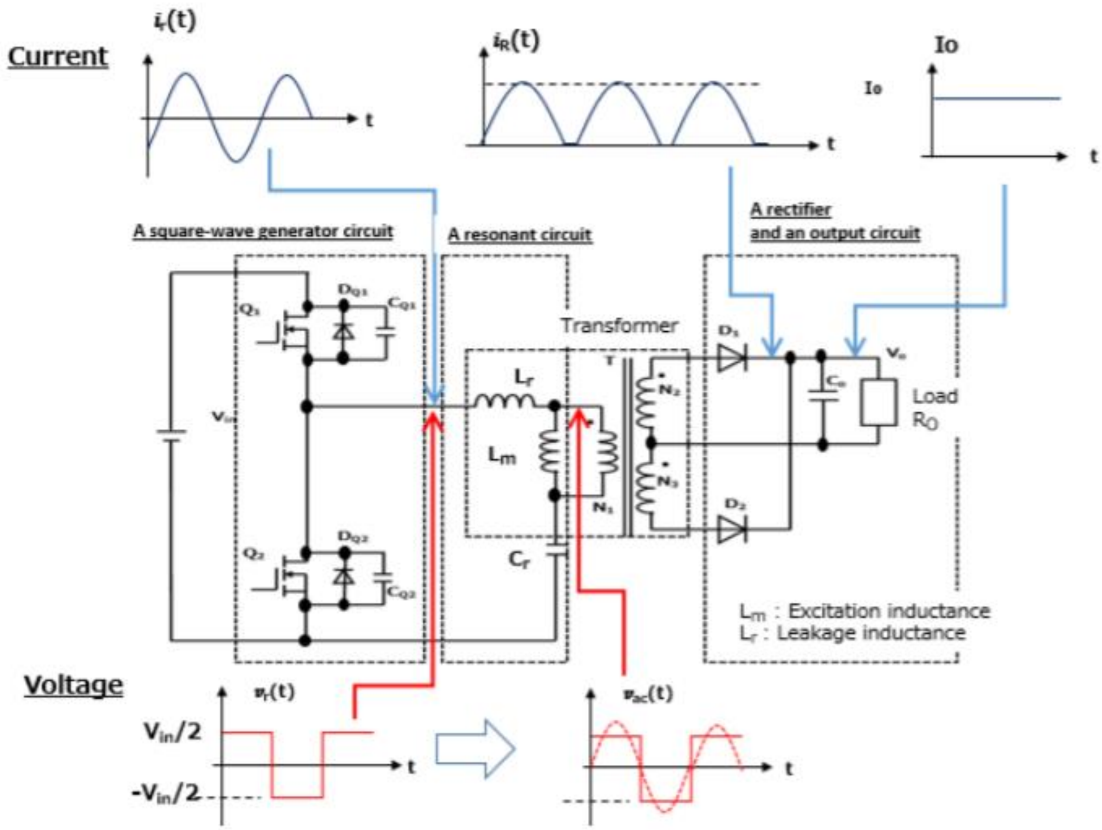


Figure 2.6 LLC resonant converter circuit

Description of the LLC resonant converter

A square-wave generator and a resonant circuit form an LLC resonant converter. An LLC resonant converter, as shown in Figure 2.6, consists basically of a) a square-wave generator circuit, b) a resonant circuit, and c) a rectifier and an output circuit.

(a)<u>**A square-wave generator circuit**</u>

Two MOSFETs (Q1 and Q2) alternately turn on and off at a frequency of f. The square-wave generator generates a square-wave voltage with an amplitude of ±Vin/2 from a DC power source with a voltage of Vin. To prevent a short-circuit between Q1 and Q2, a dead time is inserted between their on and off transitions. ZVS soft switching is performed during this period.

<u>**(b) A resonant circuit**</u>

The leakage inductance (Lr) of the transformer's primary winding, the transformer's excitation inductance (Lm) and capacitance (Cr) form a resonance circuit in sequence. A parallel load resistance circuit (Ro ') is also generated by the excitation inductance (Lm), which equivalently represents the circuit on the secondary side of the transformer as the resistance on the primary side (Figure 2.7). An LLC resonant circuit has two different resonant frequencies: a fixed resonant frequency (fr) that is a function of Lr and Cr and a resonant frequency (fm) that is determined by (Lr + Lm) and Cr. (fm varies with Ro' connected in parallel with Lm.)

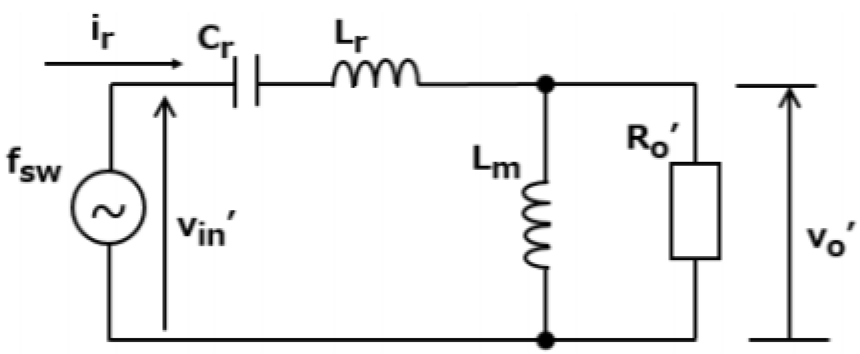


Figure 2.7 Equivalent circuit for LLC resonance

<u>**C- A rectifier and an output circuit**</u>

A full-wave rectification circuit consisting of two diodes, followed by an output capacitor (Co), is the last stage of the LLC resonant converter. Schottky barrier diodes with low forward voltage and fast reverse recovery time may be used as rectifier diodes, or MOSFETs may be used for synchronous rectification instead of diodes. Since the cathode of each diode is connected to the capacitor (voltage Vo), a square-wave voltage with an amplitude of ±Vo appears across the secondary winding of the transformer. The voltage across the primary winding also has a square waveform with an amplitude of ±N･Vo (where N = N1/N2).

2.4. Overview of the basic primary-side operation of the LLC resonant converter

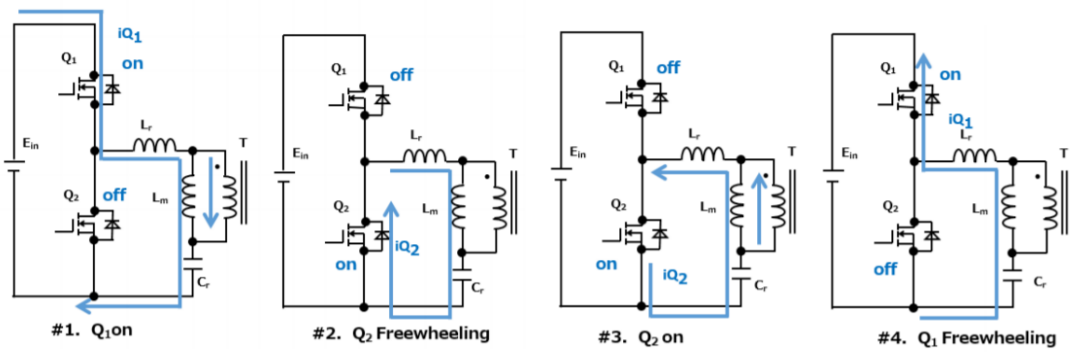


Figure 2.8 Current paths in LLC resonant converter

#1. Q1 turns on, passing iQ1.

#2. When Q1 turns off, current (iQ2) flows through the body diode of Q2 in the reverse direction. Q2 is turned on while current is flowing through the body diode.

#3. When the capacitor current (iCr) changes its direction from positive to negative because of LC resonance, iQ2 flows through Q2 in the positive direction.

#4. When Q2 is switched off when iQ2 is positive, in the reverse direction, current (iQ1) flows through the body diode of Q1. As current flows through the body diode, Q1 is switched on. At Step 2, the current is passed through the body diode of Q2, and when its voltage has reached almost zero (ZVS), Q2 is turned on. Likewise, Q1 undergoes zero-voltage switching at Step 4.

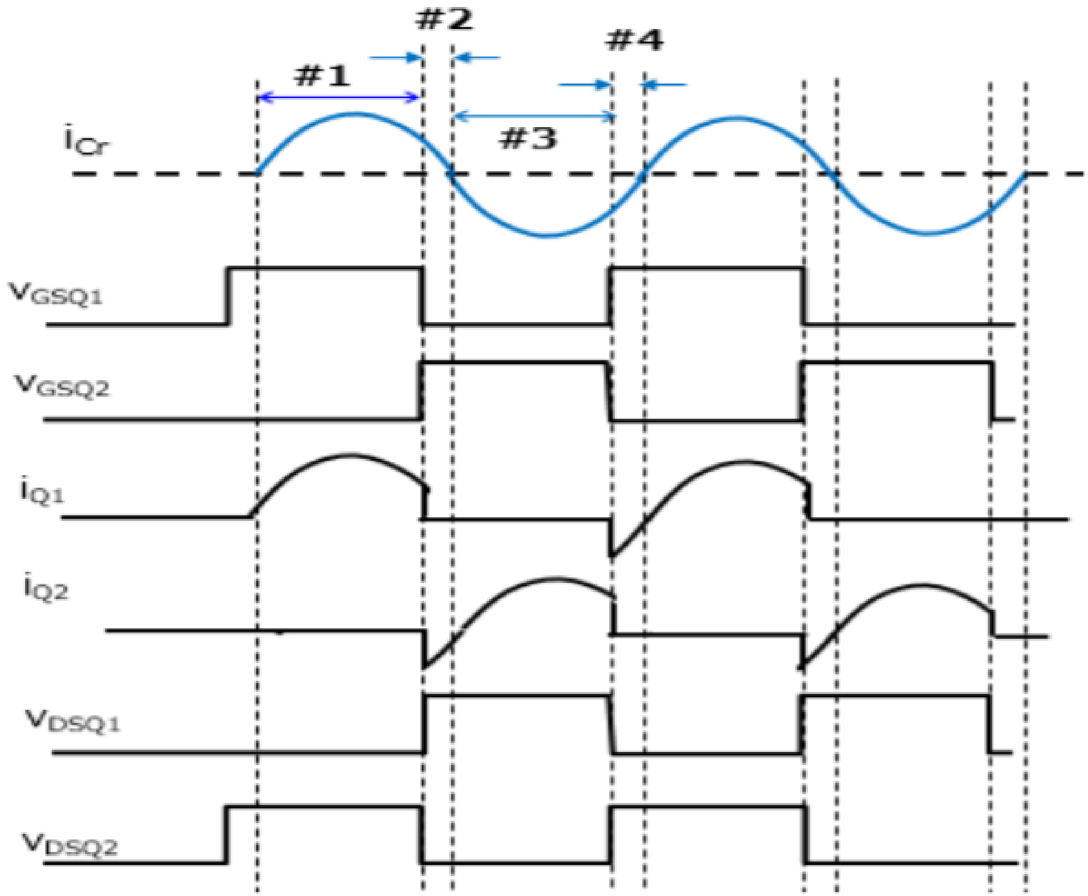


Figure 2.9 Basic waveforms of LLC resonant converter

2.5 Controlling the output voltage of the LLC resonant converter

An LLC resonant converter varies the MOSFET switching frequency (f) to regulate the output voltage by using the gain-frequency characteristics of an LLC resonant circuit, which has two distinct resonant frequencies: a fixed resonant frequency (fr) and a load-varying resonant frequency (fm). It is important to establish ZVS turn-on conditions for the MOSFETs (Q1 and Q2) to achieve the high-frequency and high-efficiency operation of the LLC resonant converter. To this end, when their drain-source voltage (vDS) is zero, the

MOSFETs are turned on. The LLC resonant converter is typically equipped with fr > f > fm to switch on the MOSFETs (Q1 and Q2) with ZVS in the square-wave generator.

For ZVS turn-off, either the MOSFET's parasitic drain-source capacitance (Coss) or an external small-value capacitor is used to reduce the VDS rise ramp of the MOSFET during turn-off and thereby turn on the MOSFET while VDS is low

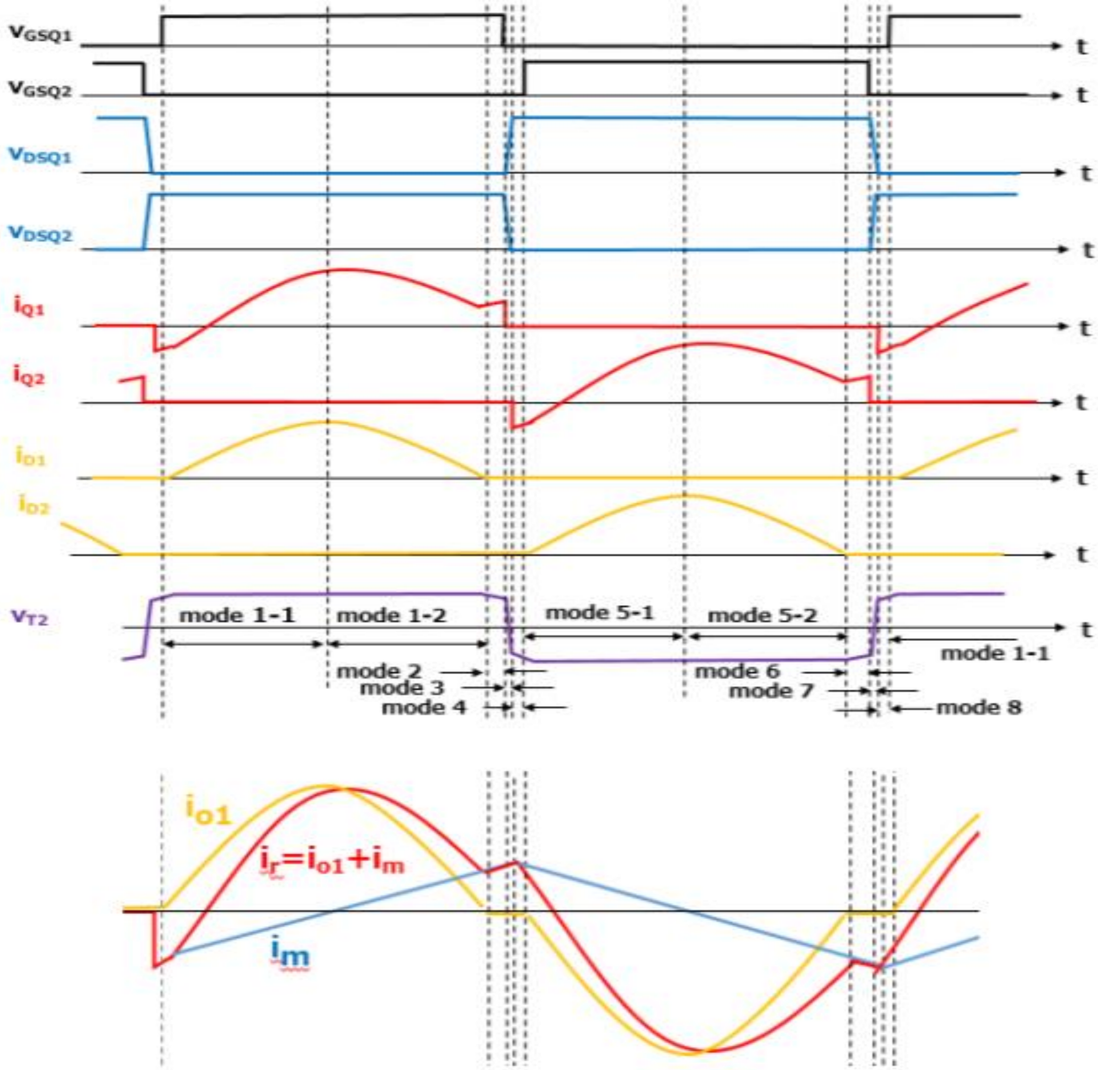


Figure 2.10 Basic waveforms of LLC resonant converter

Operating mode of the LLC resonant converter

Using frequency modulation, the LLC resonant converter regulates its output voltage. An equivalent circuit like the one shown in Figure 2.11 is used to calculate its input characteristics. (The equivalent load resistance and the actual load resistance differ.) While there is a simple topology of an LLC resonant converter, its operation is complicated. The following corresponding circuit is therefore used to estimate the resonant circuit characteristics.

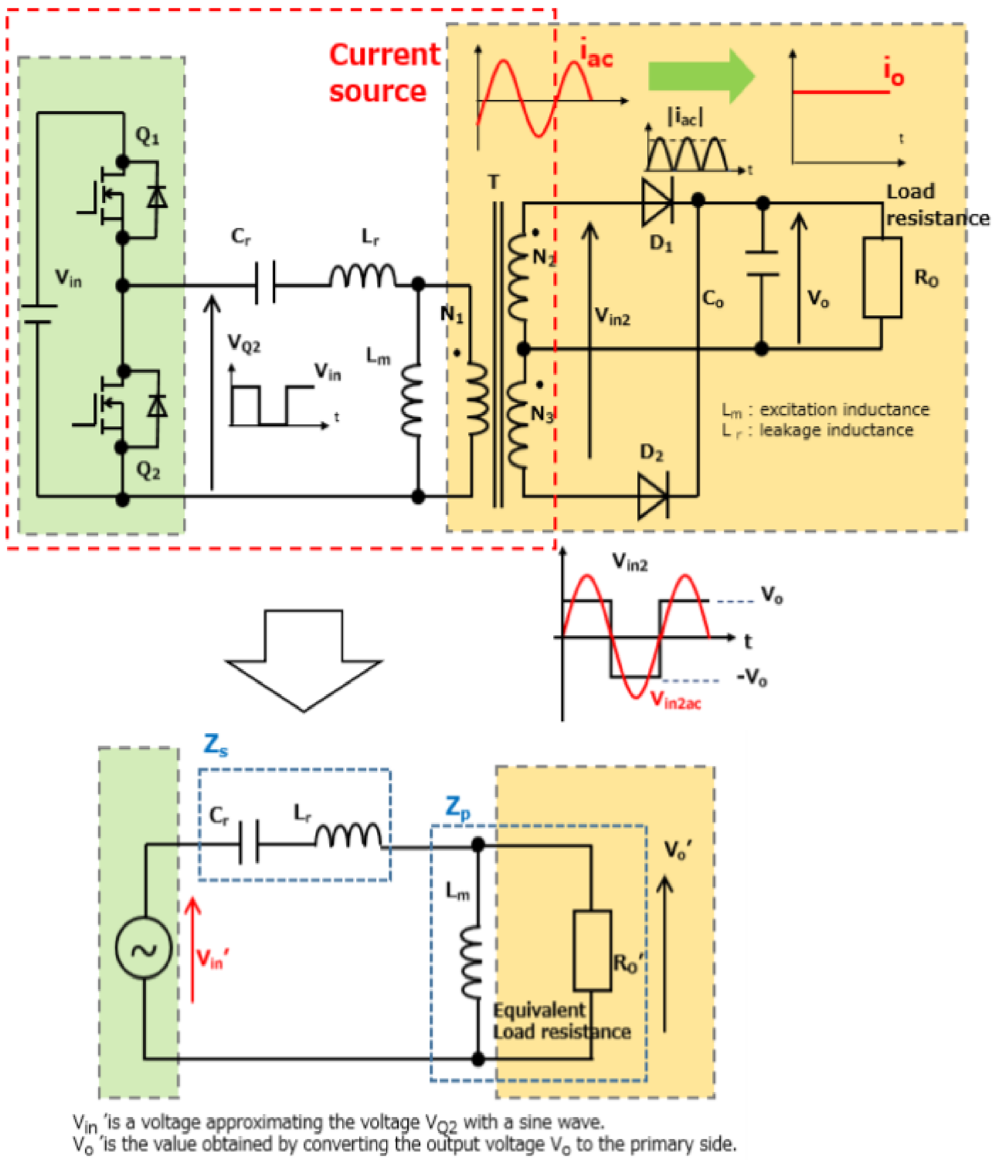


Figure 2.11 LLC resonant converter equivalent circuit

Calculating equivalent load resistance

As shown in Figure 2.11, the circuit on the primary side of an LLC resonant converter can be replaced with a sinusoidal power source (Iac). An AC voltage with a square waveform (vin2) appears at the input of the rectifier circuit on the secondary side. Because the average of |Iac| is equal to the output current (io), io and iac have the following relationship:

iac=(π・io/2)sin(ωt) (square wave to sine wave)

Vin2 is expressed as follows:

vin2=+vo (when sin(ωt) > 0)

vin2=-vo (when sin(ωt) < 0) (where vo is the output voltage.)

The fundamental (sine-wave) component of vin2 (vin2ac) is expressed as follows: (The harmonic content of vin2 has no effect on power transfer.)

vin2ac=(4vo/π)sin(ωt)

The AC equivalent load resistance (Ro') observed on the primary side can be calculated as follows from the division of vin2ac by iac and the turns ratios of the transformer.

Ro'=n2 ×vin2ac/iac

Ro'=(8n2 /π2)・Ro

The ratio of transformer winding voltage is directly proportional to the ratio of winding turns, while the ratio of winding current is inversely proportional to the ratio of winding turns. Resistance is voltage divided by current and proportional to the square of the ratio of winding turns.

Input/output voltage ratio

(vin' is a sinusoidal approximation of Vin. vo' is obtained by converting vin2ac approximating sine wave of Vo to the primary side.)

The relationship between vin' and vo' in Figure 2.11 can be calculated with a complex-valued function as follows:

$$Z_s = \frac{1}{j\omega C_r} + j\omega L_r \ , \ Z_p = \frac{1}{\frac{1}{j\omega L_m} + \frac{1}{R_0^2}}$$

$$\frac{v_0^2}{v_{in}^2} = \frac{Z_p}{Z_s + Z_p} = \frac{1}{1 + \frac{Z_S}{Z_P}}$$

$$= \frac{1}{1+\left(\frac{1}{j\omega C_r}+j\omega L_r\right)\left(\frac{1}{j\omega L_m}+\frac{1}{R_0^2}\right)}$$

$$= \frac{1}{1 + \frac{L_r}{L_m} - \frac{1}{\omega^2 L_m C_r} + j\left(\frac{\omega L_r}{R_0^2} - \frac{1}{\omega C_r R_0^2}\right)}$$

Because

$$\omega = 2\pi f\ (f: switchnig\ frequency), \omega_0 = \frac{1}{\sqrt{L_r C_r}}, Q = \frac{1}{R_o'}\sqrt{\frac{L_r}{C_r}}$$

The above equation can be restated as:

$$\frac{v_o'}{v_{in}'} = \frac{1}{1 + \frac{L_r}{L_m}\left(1 - \frac{\omega_0{}^2}{\omega^2}\right) + jQ\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right)}$$

Hence,

$$\frac{|v_o'|}{|v_{in}'|} = \frac{1}{\sqrt{\left(1 + \frac{L_r}{L_m}\left(1 - \frac{\omega_0{}^2}{\omega^2}\right)\right)^2 + Q^2\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right)^2}}$$

An LLC resonant converter has different operating modes above and below the maximum value of the input/output voltage ratio. The LLC resonant converter exhibits a capacitive operation in the region below the frequency at which it provides the maximum input/output voltage ratio. In this frequency region, the upper- and lower-arm MOSFETs are short-circuited. Generally, a capacitive operation is prevented by operating the LLC resonant converter at a higher frequency. The LLC resonant converter is not also used in this frequency region because of poor controllability since the output voltage changes only slightly in response to changes in f in the region where the switching frequency (f) is higher than the resonance frequency (fr). Figure 2.12 shows the relationships between the input/output voltage ratio and the switching frequency.

1. When f=fr, Cr and Lr are in series resonance. In this state, Lr has zero impedance, causing Vo' to be equal to Vin'. Therefore, the output voltage does not change in response to changes in the load resistance (RO).

2. When fm<f<fr, the out put voltage decreases under heavy load with lower Ro'

3. When f approaches fm, the output voltage increases. This is because, the Lm voltage, which increases as a result of series resonance between Cr and Lm＋Lr, is applied to the transformer.

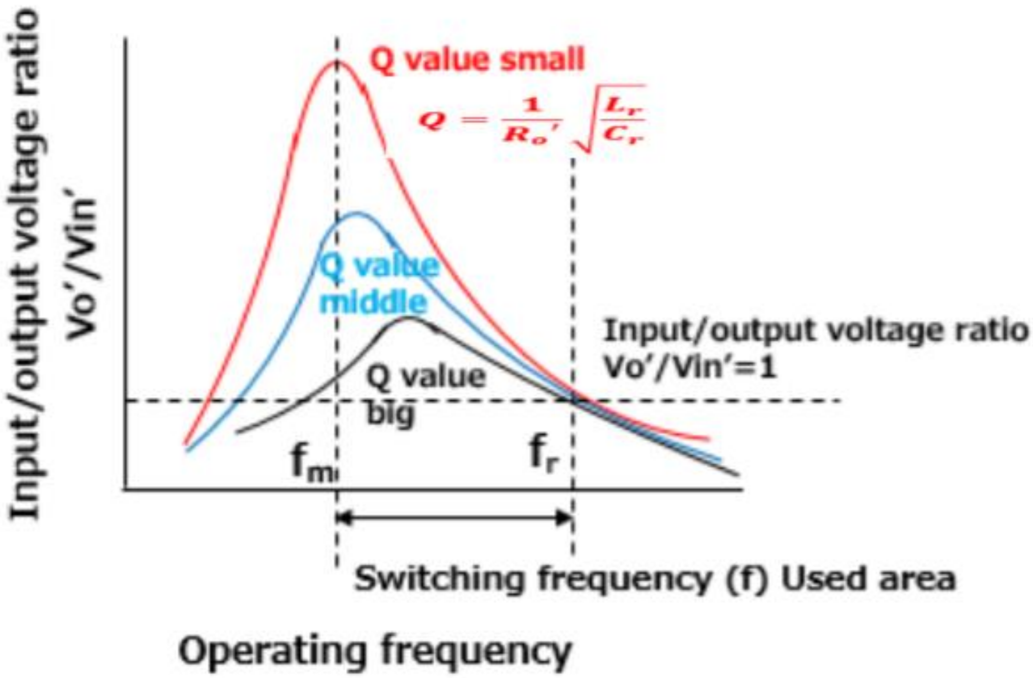


Figure 2.12 Output voltage curves

Primary-side current of LLC resonant converter

The current on the primary side of the LLC circuit consists of a resonance current flowing through Lr and Cr and a resonance current flowing through Lr+Lm and Cr.

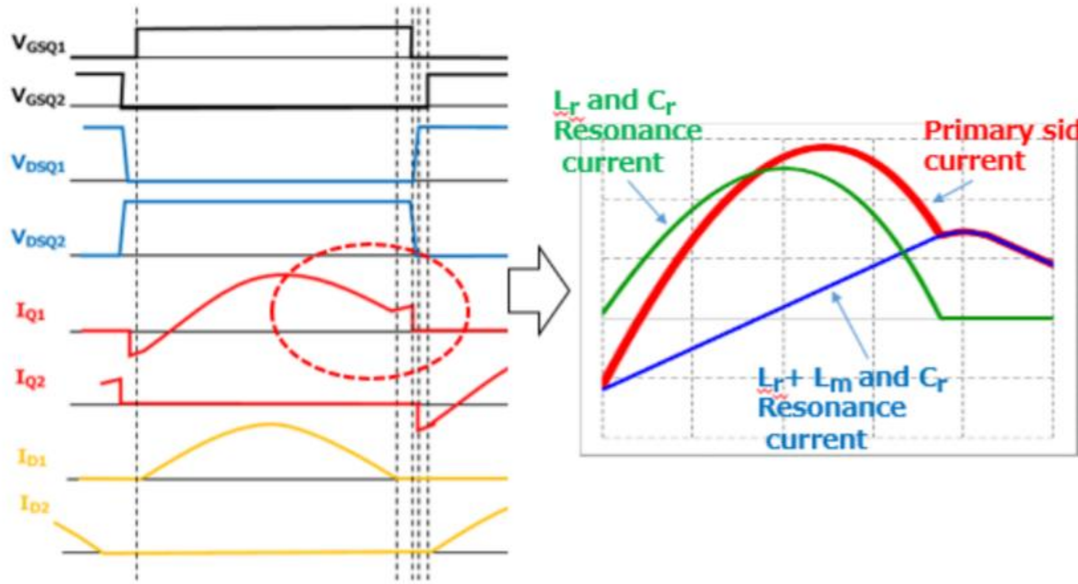


Figure 2.13 Waveforms of the primary-side current of the LLC resonant converter

Frequency and load of the LLC resonant converter

When the MOSFET switching frequency (f) is equal to the resonant frequency (fr)

Figure 2.14 is a simplified version of the equivalent circuit shown in Figure 2.6. In this circuit, the input can be considered to be short-circuited with the series resonant circuit (consisting of Lr and Cr) because they have the same frequency. The output voltage (vo') across the load resistor (Ro') is approximately equal to vin'. Therefore, the input/output voltage gain (η) is vo'/vin' ≈1 regardless of the relationship of the load Ro' and fm (i.e., the resonant frequency of Lr+Lm and Cr).

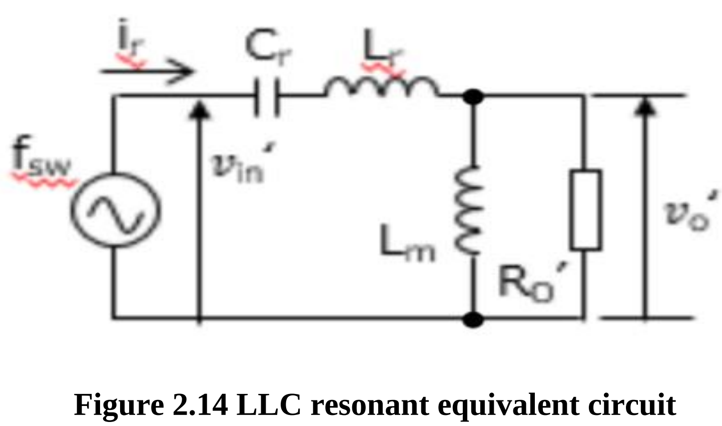


Figure 2.14 LLC resonant equivalent circuit

When the load (Ro') is heavy (i.e., when Ro' is small and the load current Io is large)

When ωLm >>Ro', the parallel circuit of Lm and Ro' can be considered to consist of only resistor Ro' as shown in Figure 2.15. Therefore, the resonant frequency (fm) is close to fr that is determined by the series resonance between Lr and Cr. In this case, the input/output voltage gain (η=vo'/vin') remains less than 1

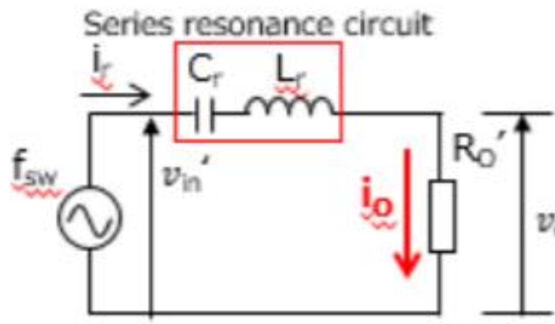


Figure 2.15 Resonant circuit under heavy load

When the load (Ro') is light (i.e., when Ro' is large and the load current io is small)

When ωLm << Ro', the parallel circuit of Lm and Ro' can be considered to consist of only inductor Lm as shown in Figure 2.16. Therefore, (Lr+Lm) and Cr form a series resonance circuit. The resonant frequency (Fm) deviates toward a lower value from Fr of the series resonance circuit consisting of Lr and Cr. When Ro' is open (i.e., Ro'= ∞)

voltage gain (η=vo'/vin') becomes the maximum

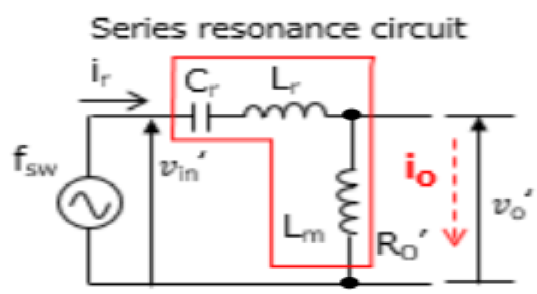


Figure 2.16 Resonant circuit under light load (Ro')

A resonant circuit generates an almost sinusoidal current from a square-wave voltage and feeds it to a rectifier circuit on the secondary side. The output voltage (vo') can be changed by changing the switching frequency of the square wave generator.

2.6 Examples on Applications using resonant converters

2.6.1 Resonant Converter for Photovoltaic Applications

As compared to more traditional dc-dc converter configurations, Power conversion for photovoltaic (PV) applications requires and adaptable device that can respond to a wide range of input voltage and current conditions.

If a converter is only configured for high peak performance, the range of conditions common to many PV installations sometimes cause the converter to be much less effective in another operating area. The PV voltage varies considerably with the panel construction and operating temperature, while the PV current changes largely due to solar irradiance and shading conditions. In the PV power conditioning system (PCS) design process between the PV panel and the electrical utility system, the importance of galvanic isolation is essential. It helps to increase the voltage boost ratio, decrease the current of ground leakage and overall protection during conditions of fault. As several authors have already proposed, distributed maximum power point tracking (MPPT) can achieve much better energy harvesting over systems that are completely centralized

Researchers have also concluded that a system structure with the PV panels connected in parallel can be much more productive in low-light and partially shaded conditions than a series-connected system.

Block diagrams showing the micro inverter and micro converter system structures are provided in Fig. 2.17

This combination of high performance, galvanic isolation and a localized, distributed approach to the conversion of energy has prompted technological progress to continue.

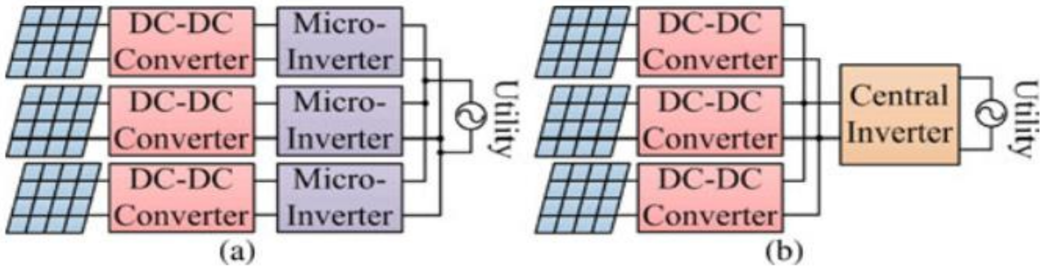


Figure 2.17 Distributed (a) microinverter and (b) micro converter system structures

There have been a variety of different methods proposed for micro conversion:

- Isolated
- Non-Isolated

Some utilize one pulse width-modulated (PWM) stage in the conversion from dc– ac, others utilize multiple PWM stages.

Multi-stage solutions allow for greater interoperability between distributed AC and DC systems, thus enabling the elimination of short-lived electrolytic capacitors. The isolated dc-dc stage in the distributed PV PCS must operate effectively at full power, while maintaining high performance over a PV voltage range at light load. It is essential to minimize the amount of circulating energy in the system to maintain high performance under low-power conditions. The production of a system with a high-power factor at the isolation transformer would be an alternative definition of this characteristic.

The series-resonant converter and, more recently, the LLC resonant converters are an option, both operating on a similar principle and typically using a variable frequency control to adjust the output voltage. When the series-resonant, or LLC converter, is operated near the resonant frequency of the tank circuit, the converter achieves almost zero-voltage switching (ZVS) and zero-current switching (ZCS) with very low circulating energy, giving it a high peak efficiency. However, as the operating frequency diverges from the resonant frequency, the sum of circulating energy increases. Unfortunately, the normal conditions for PV conversion will often push the converter significantly away from the optimum switching frequency.

Several authors have proposed methods to extend the line and load range of the LLC, once again complicating the circuit topology and control .

Other authors have proposed using the series-resonant converter as an unregulated dc–dc transformer (DCX).

This approach has the benefit of almost no switching loss, little or no circulating energy, very high peak efficiency, and integrated isolation.

This concept of using the series-resonant DCX is not without merit, but the system requires an additional element to provide regulation capability.

This idea of using the series-resonant DCX is not without merit, but in order to provide regulation functionality, the device needs an additional feature.

2.6.2 High-Frequency Unregulated LLC Resonant Converter for Fuel Cell Applications.

The fuel cell is one of the most powerful energy sources among many forms of renewable energy sources that are promising in the near future. Electrically driven devices, such as transportation, communication, computing, and residential systems, may be an alternate energy source. However under variable load conditions, the fuel cell shows a wide variance in output voltage, and the voltage provided by the fuel cell is low in magnitude, so that the voltage and current in applications with the output power of the kilowatt range can easily have the same order of magnitude.

Moreover, for power conditioning system design, the low fuel cell voltage should be elevated to the height of a utility line voltage.

In order to achieve a high voltage gain, converters based on a transformer or coupled inductor have been considered. Compared with an isolation transformer, the coupled inductor has a simple and efficient structure, but its use is limited to applications that do not need electrical isolation. In order to achieve not only high voltage gain, but also electrical separation, current-fed converters have been commonly used. Figure 2.18(a) shows a two-inductor current-fed half-bridge converter. The significant drawbacks of this converter are hard-switching operation and turnoff voltage spikes. To alleviate these issues, a zero-voltage switching (ZVS) approach using an active-clamp circuit can be used, but this active-clamp circuit increases the complexity of the circuit due to additional switch blocks. Using the current-fed full bridge converter with two magnetic devices, as is shown in Figure 2.18(b), the number of magnetic elements that influence the circuit volume can be decreased, but the switch blocks are increased to five, including the active-clamp-circuit.

Using the current-fed push-pull converter, shown in Fig 2.18(c), the circuit structure can be simplified. But it does have several drawbacks, such as the switches' high voltage stress and low conversion of voltage ratio. In general, the switching frequency of the current fed converters can be limited to under several tens of kilohertz due to switching losses. At a 1 kW configuration (Vin = 22 V ~ 41 V/Vo = 350 V) with a 100 kHz switching frequency, they are not stated to be able to overcome the efficiency of 89% Therefore the additional circuitry to alleviate the voltage spikes and the performance deterioration due to the switching loss can be barriers to reducing the size and expense of the current-fed converters due to the transformer leakage inductance. Another notion of an isolated dc/dc converter is proposed for fuel cell applications. This converter consists of two stages, namely the hard-switched output voltage control boost converter and the unregulated electrical isolation and voltage amplification LLC converter. The unregulated LLC converter can be configured to have an ideal switching characteristic, regardless of the load conditions, by separating the functions so that the switching frequency can be increased to 300 Khz.

With 1-kW (Vin = 24 V ~ 48 V/Vo = 400 V), the feasibility of the proposed converter has been verified.

It is composed of the output voltage control boost converter and the isolated voltage amplifier using the unregulated LLC converter. From the experimental results from the 1-kW prototype converter that has been designed with switching frequencies of 50 kHz in the boost stage and 300 kHz in the *LLC* stage, an efficiency of more than 90.2% has been obtained at a 24-V input

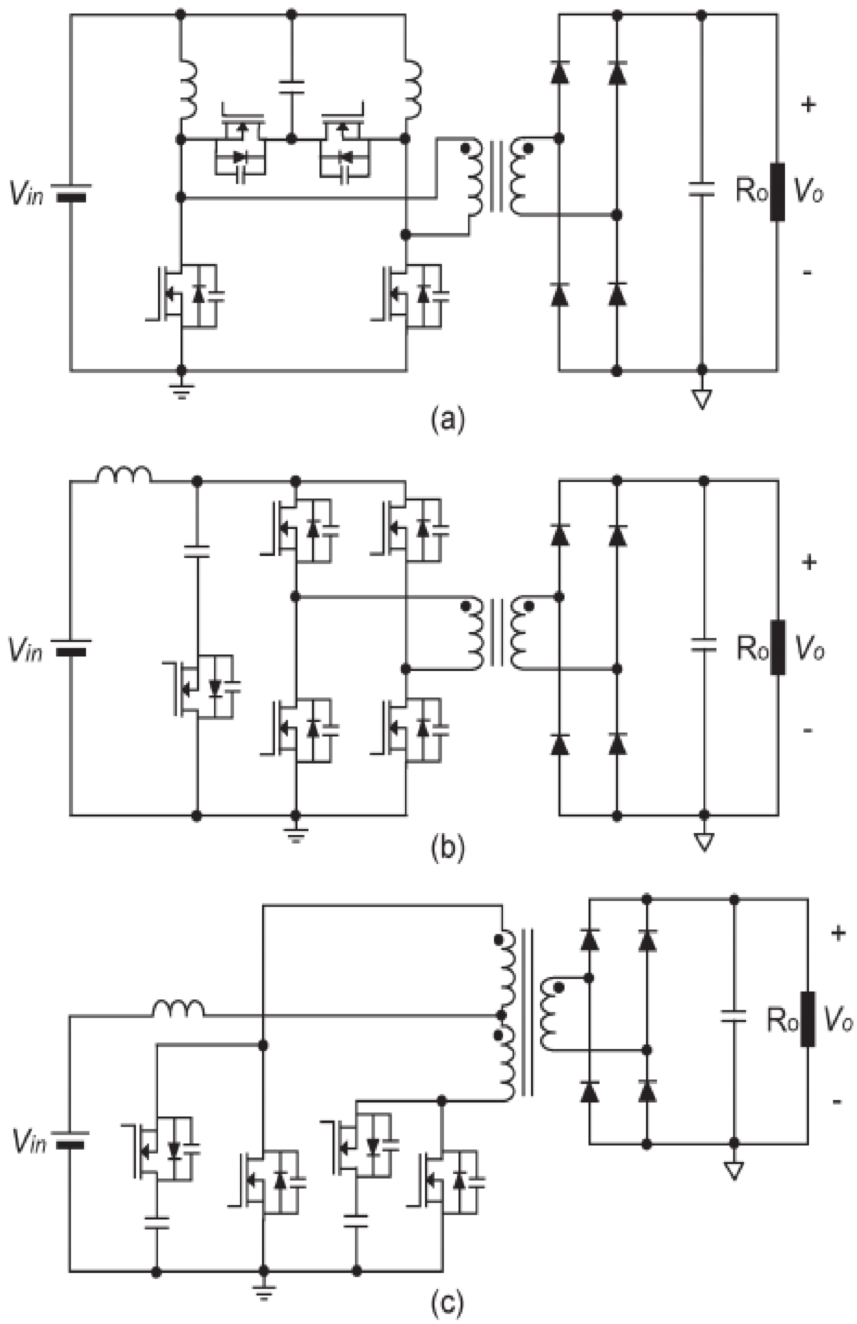


Figure 2.18. Current fed type converters with active clamp circuit. (a) Half-bridge type (b) Full-bridge type. (c) Push-pull type.

2.6.3 Bidirectional DC- DC Resonant Converter for Vehicle to-Grid (V2G) Applications

Electric vehicles (EVs) and plug-in hybrid electric vehicles (PHEVs) have gained popularity because of their advantages, such as more environmentally friendly, less noisy and more efficient.

In vehicle-to-grid (V2G) applications, battery chargers are bidirectional, capable of transferring power back to the grid.

Vehicles with V2G capability require a bidirectional battery charger that consists of a bidirectional ac-dc converter followed by a bidirectional dc-dc converter. The dc-dc converters manage the power flow between the dc bus and the battery.

The main target in these converters is

- Achieve soft switching in the power switches for a wide range of load variation.
- Enable very high frequency operation

So that the size and the cost of the output filter is reduced. Both Isolated and non-isolated converters are considered.

The advantages with the non-isolated converters are fewer numbers of components and high power-stage efficiency. However, in some applications, including EV battery chargers, galvanic isolation is required for safety between the primary-side and the secondary side.

The ZVS range depends on the primary-side transformer leakage inductance. To achieve ZVS for all load and line conditions, large leakage inductance is required. This, however, results in duty cycle loss and it increases the reactive energy in the circuit for certain load conditions, which decreases converter efficiency.

Due to the energy contained in the secondary-side leakage inductance of the transformer, the secondary side switches have extreme parasitic ringing. RCD snubbers, lossless snubbers or active clamp circuits are used to dample or clamp this ringing. These additional circuitries cause the voltage spikes across the secondary switches to be reduced. However, they increase the complexity of the size, cost and circuit. Also with these snubber or clamp circuits, the voltage stress on the secondary side switches is greater than the output voltage, rendering this topology unacceptable for applications with high output voltages. For high-power dc-dc applications, the optimal topology is resonant converters.

In some applications, resonant converters are the obvious option because the reactive effects are noticeable and inevitable, whereas in other applications, these converters are chosen because they provide a variety of advantages over non-resonant type topologies, including soft-switching in all switches, very high frequency operation, low EMI, easy control, high performance, and less component counts. It removes the need for clamps or snubbers for circuitry. The LLC converter is one of the most common resonant topologies for a wide input and large output voltage range. This topology, however, is ideal for unidirectional power transfer.

The magnetizing inductance in the regeneration mode is parallel to the bridge voltage, and it is no longer part of the resonant network, converting the topology into a series resonant converter (SRC). The efficiency of the SRC drops drastically as the operating switching frequency drifts away from the series resonant frequency, making it unsuitable for applications with a very large voltage range of input and output.

The CLLLC model of resonant converter is identical to the LLC type of resonant converter with an additional inductor and a secondary side capacitor. In all the switches without additional circuitry, soft switching can be ensured. Because of soft switching in all switches, very high-frequency operation is possible, so the size of the magnetics and the capacitors can be small. To minimize the size and cost of the converter, a CLLC-type resonant network is derived from the initial CLLLC-type resonant network.

It was found that the CLLC-type network was more compact and lighter in weight as compared to the CLLLC-type network. For a CLLLC configuration, the peak efficiency was 97.7% under BCM and 98.1% under RM. For a CLLC configuration, the peak efficiency was 97.7% under BCM and 97.9% under RM.

2.6.4 Resonant converter in Ozone generator model

Nowadays, ozone is becoming the oxidant of choice for many air and water applications. As the oxidant element for bleaching and disinfecting, ozone is widely used in industrial and domestic applications. It is also used in food processing, food storage, odor reduction, groundwater remediation, and drinking water purification. Ozone use is a rapidly growing field in which improvements in ozone generation systems are a key issue.

As opposed to the traditionally used low-frequency power supplies, the use of high-frequency converters to supply the Ozonizer is the first option to increase efficiency. High frequency converters provide lower power losses, lower size and weight and the possibility of controlling the amount of ozone generated. The basic ozone generation configuration is shown in Figure 2.19. The ozone generator (OG) principle of operation consists of applying a high voltage between two parallel plaques with air inside, producing a high voltage phenomenon known as a silent discharge or corona effect. The silent discharge generates ultraviolet radiation that breaks down ozone-producing oxygen molecules. To maintain the silent discharge, which is obtained by adding a dielectric between the air gap and one of the electrodes, it is important to avoid arc discharge.

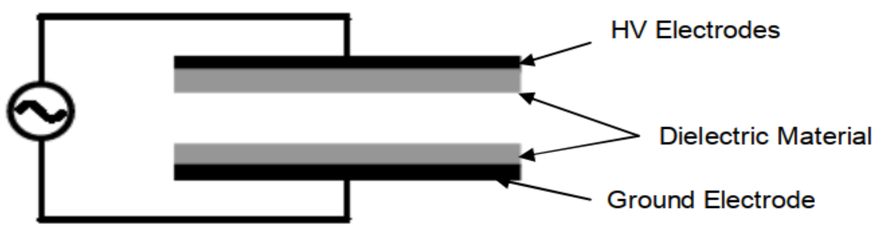
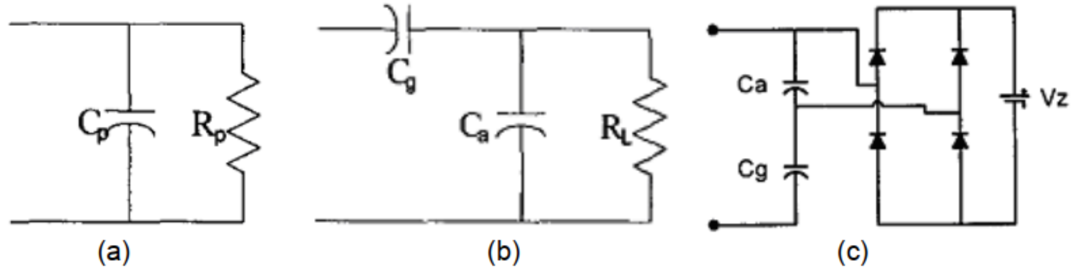


Figure 2.19 The basic configuration of ozone generation



The most used model for designers is shown in Figure 2.20

Figure 2.20 The most used model of the OG
a. Linier model of ozone generator which consists of capacitor and resistor
b. Linier model of ozone generator which represent capacitance of air gap and dielectric
c. Non linier model of Ozone generator

Figure 2.20.a. Ozone generators are a linear model. With only one resistor and capacitor, this model represents the OG. The way they achieve this linear model is based on the analysis of power.

Figure 2.20.b. It is also an OG liner model that takes into account that capacitance is represented by Cg due to the dielectric and that capacitance is represented by Ca due to the air gap.

Figure 2.20.c. Representing the non-linear OG model.

The DC source with the diode bridge represents the silent discharge, where Vz is the voltage at which the silent discharge is generated. It represents the non-liner model of OG. This is a non-linear model, because when the applied voltage across Ca is lower than Vz, and the bridge diodes do not conduct, there is no silent discharge. On the other hand, when the voltage applied across Ca is greater than Vz, the diodes conduct, and the applied voltage is equal to Vz, the silent discharge is shown, and the effect of Ca disappears. Due to this non-linear characteristic of Fig 2.20.c model, it is difficult to design the power supply basing on this model.

For low power application the using of half bridge inverter, has been well recognized as power converter and power supply. Due to its high efficiency and simple circuit this model has been used in other application such as in **electronic ballast**.

The use of a simple resonant inverter supplied by the ozone chamber model 2.20.a is proposed.

In order to determine the operating frequency of the ozone generation, the simplified resonant RLC circuit has provided good practice. The use of resonance inverters for ozone chamber power supply has been discussed. The best result of the ozone generator resonance inverter will be obtained if the inverter supplies the electrical voltage at the same load or ozone chamber frequency.

3. High Frequency Transformers

3.1 What is a Transformer?

A transformer is an electrical device that transfers energies between two or more circuits by electromagnetic induction. A varying current in the primary winding of a transformer creates a varying magnetic flux in the core and a collision of the secondary winding with a varying magnetic field. A varying electromotive field (emf) or voltage in the secondary winding is induced by this varying magnetic field in the secondary.

Maximum Power in a Compact Package

Figure 3.1 Standard Transformer

The electronics of today require significant transformers that fit into the most compact applications .

Any piece of equipment that runs on poorly designed transformers runs the risk of breakdowns and failures. The need for high frequency transformers continues to grow as digital electronics become part of an ever-increasing number of devices.

The following list represent some applications that are associated with these transformers:

- Personal electronics
- PCs
- Industrial equipment
- Solar converters
- Electric drives
- Energy conversion

Basic Principles

High-frequency transformers operate using the same basic principles as standard transformers. The essential distinction is that, as their name suggests, they work at a lot higher frequencies — while most line voltage transformers work at 50 or 60 Hz, high-recurrence transformers use frequencies from 20 KHz to over 1MHz. There are several advantages to working at a greater frequency, the first of which is size of the transformer. The greater the frequency, the smaller the transformer can be, for any given power rating. Second, less copper wire is required because the transformer is smaller, thereby reducing the losses and helping to make the transformer more efficient. A broad range of geometries are also available, since the core is usually ferrite, so that the transformer can be customized for the application. The chances are good that a ferrite core exists to meet the requirement, whether additional shielding or a particular form factor is needed. However, the benefits brought about by light weight, small size, and higher power density, pose several challenges. In the design of high frequency transformers, reducing problems such as skin and proximity effects is a serious concern.

3.2. Transformer Basics

3.2.1. Geometry of Copper Windings and Core Wire Winding Window

Let 's consider only two wire windings wound on one magnetic core for simplicity, which acts to couple the magnetic flux with near unity transfer between the two coils.

In Switch Mode Power Supplies, the main objective of a power transformer is to transfer power from an external electrical source to external loads placed on the output windings efficiently and instantaneously. In doing the following, the transformer also provides important additional capabilities:

- To achieve multiple outputs at different voltage levels, multiple secondaries with different numbers of turns can be used.
- Turns ratio of primary to secondary can be maintained to effectively accommodate widely several input/output voltage levels.
- Separate primary and secondary windings, particularly important for safety, facilitate high voltage input/output isolation.
- For an optimum transformer winding partition, a set of k wire windings brings some complexity to the issues. For the case of multiple windings, the competing issues are more Complex.

All the wire windings wound on a given core must fit into its one wire winding window which we term either A_W or W_A in the text below.

In the literature of transformers, both symbols are found. The voids between round wires, wire insulation and any bobbin structure on which the wire turns are mounted, as well as insulation between high voltage and low voltage windings, take up much of the actual winding area.

Practically, only about 50% of the window area can carry an active conductor. “Fill factor” is what this fraction is called. In a two-winding transformer, this means that each winding can fill not more than 25% of the total wire winding window area

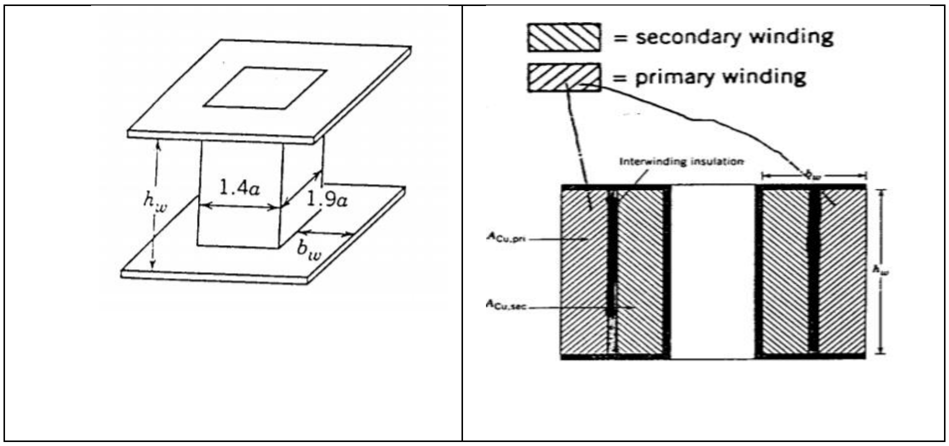


Figure 3.2 Area for windings

-Total area for windings: A_W = Apri + Asec

* According to the required wire sizes (AWG #) in each coil, Aw is split into two parts for a two winding transformer, which is in turn chosen for the expected current flow to prevent overheating of the wires.

The Primary wire winding area occupied in the wire winding window

$$A_w\,(\text{prim}) = \frac{N_{pri}\,A_{cu\ wire}\ (\text{AWG \# of primary})}{K_{cu}\,(\text{primary})}$$

With different wire type choices for the primary and secondary K cu (primary) Kcu (secondary). If we want heat from I^2R or pJ^2 losses to be uniformly distributed over core window volume, then For primary & secondary windings, we usually choose the same type of wire to obtain similar Kcu, but AWG# varies to achieve the same J in the two winding sets. In That way both windings will heat uniformly.

if -> pcuJ^2(primary) = pcuJ^2(secondary)

$$\uparrow \qquad\qquad \uparrow$$

$$\left(\frac{I_{prim}}{A_{cu}(\text{primary})}\right)^2 \qquad \left(\frac{I_{sec}}{A_{cu}(\text{secondary})}\right)^2$$

Since both wire windings must fit in the core wire window area Aw or WA we have a limitation : In short, the wire size is chosen to support both the desired current level and to fit into the window area of the core. Aw(window for wire)= A(prim) + A(sec) The largest wire that will fit into the window is often used in low frequency windings. This minimizes losses and maximizes a transformer 's power rate.

3.2.2 Single Wire Skin Effect multi-wire and proximity effects.

The tendency of high frequency currents to flow on the surface of conductors causes **skin effects**. The tendency to go to the wire surface for high frequency currents in wire coils changes the effective A(wire) in turn causing the expected Cu loss increases from windings at high frequencies compared to low frequencies.

Proximity effects increases wire losses via MMF effects and this is due to the collective magnetic field from many wires.

Proximity effects are also recognized as losses of eddy current are caused either by magnetic fields from adjacent conductors in adjacent windings or, more seriously, in adjacent layers, causing current to flow in unintended patterns or eddy currents. This effect creates excessive resistance and unintentional power loss within the wire. There are a wide range of design considerations that minimize proximity effects, one way is to select a core that allows an increased number of turns / layers. Like interleaving the winding, the use of foil winding layers rather than round wire is another. Skin effects in all wires cause non-uniform current density profiles, but non-interleaved windings enable mmf to be built in the wire winding window.

For each wire position, this mmf is distinct and ruins the assumption that J is distributed the same way for all wires uniformly across the wire area. That is, various wires in the real world of hf transformers have different J distributions.

According to its position in the wire winding window, the increasing MMF spatial variation makes the R (each wire turn) unique in its losses. The more non-uniform the J, the higher the mmf seen by the wire.

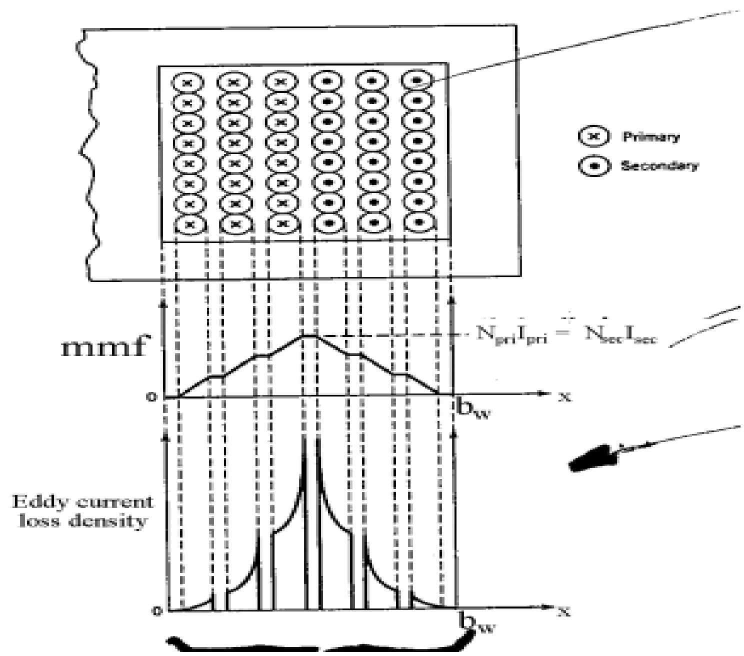


Figure 3.3 J Flow on the surface only would cause more Cu losses

Ideally, a transformer does not store energy, but all energy is instantaneously transferred from input to output coils. In practice, compared to ideal transformers that have no inductances associated with them, all transformers store some energy in the two types of inductances associated with the real transformer.

There are two inductances:

- Leakage inductance, L*i*, represents energy stored in the wire winding windows and area between wire windings, caused by imperfect flux coupling for a core with finite magnetic flux. The leakage inductance in the equivalent electrical circuit is in series with the wire windings. And the energy stored is squared in proportion to the wire winding current.
- Magnetizing inductance, LM , represents energy stored in the magnetic core and in small air gaps that occur when a closed magnetic loop core is formed by the separate core halves. Mutual inductance only occurs in parallel with the primary windings in the equivalent circuit of a real transformer. The energy contained in

the inductance of magnetization is a function of the volt-seconds applied to the primary windings per turn and is independent of the load current.

3.2.3. Various Frequencies of Interest to Magnetic Devices and Copper Loss

There are plenty of meanings to the term "Frequency" in switching applications.

- "Clock Frequency" is the frequency of the clock pulses generated in the control integrated circuit.
- "Switching Frequency" is usually defined as the frequency at which switch drive signals are generated. It is also often the frequency seen by the output filter, the frequency of the output ripple and the current of the input ripple and is an important concept in the design of the control loop. The power switch, the transformer, and the output rectifier all operate at the switching frequency without confusion in a single-ended power circuit, such as the forward converter.

3. 2. 4. Transformer Equivalent Circuit

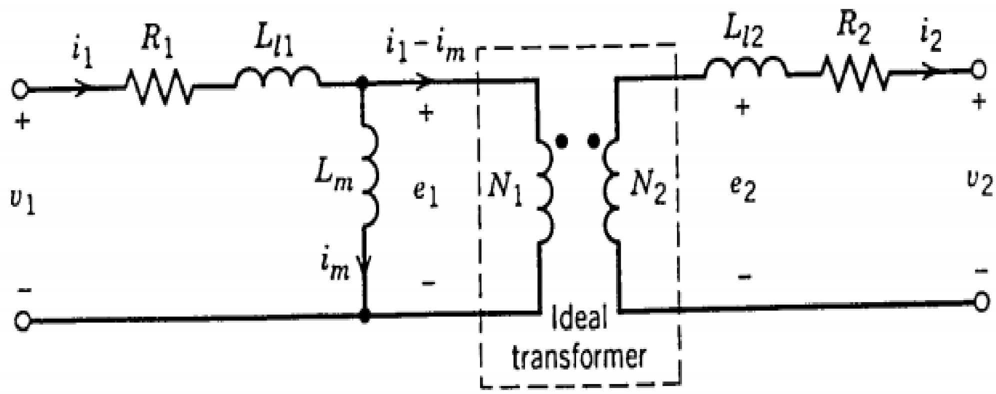


Figure 3.4 Equivalent Circuit of Transformer

- R1 = R2 = 0 only for superconducting wire, but will be larger than the expected DC wire resistances for all AC currents due to both skin effects and proximity effects
- For μc(core) → ∞ then ℜ c = 0 ⇒ Lm → ∞ The big magnetizing inductor draws no current from the input voltage. This means ⇒ Im is negligible compared to i1. Moreover, ℜ *l*1 and ℜ *l*2 → ∞ ⇒ L*l* → 0 because leakage flux ≡ 0. This is not the case in practice. Consider the extreme case of the fly back "transformer",

where we purposefully make Lm to be small so we can store lots of energy in the air gap.

- The magnetizing inductance is solely a core 's magnetic property and is not effected at all by the load current drawn by the transformer secondary. It's simply shown by:

$$L_m = \frac{N_1^2}{\Re_c}$$

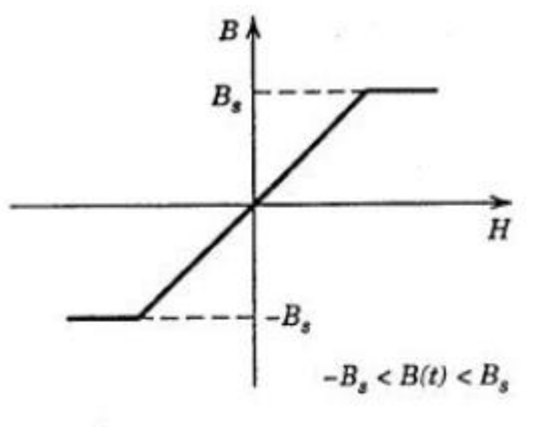
That is, the net MMF from the copper windings N1i1 + N2i2 is supposed to be small, almost zero for operating transformers and there are no amp-turn constraints The voltage placed across the transformer's coil is, however, subject to volt-sec limitations. That is above a critical current level, the above magnetizing inductance will change subject to the level of the induced magnetizing current.

IM levels are driven only by input volt seconds conditions, which determine the magnetizing current under AC drive conditions.

H(core) l(core) = n im

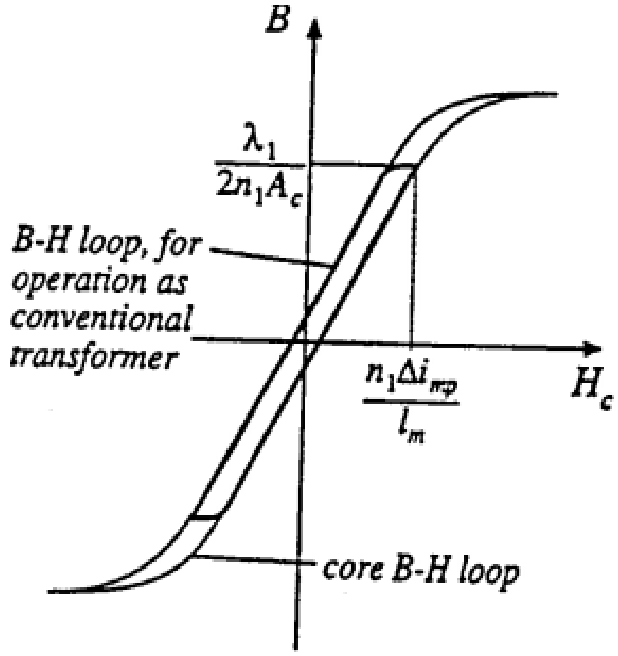
If n im is too high, then H exceeds H(critical) or BSAT , Then the core saturates causing $\mu r \rightarrow \mu o$.

To determine B(max) for a transformer driven by AC signals we employ the flux linkage, $\lambda = N\phi$, and $VL = d\lambda/dt$. This gives a volt-sec limit.



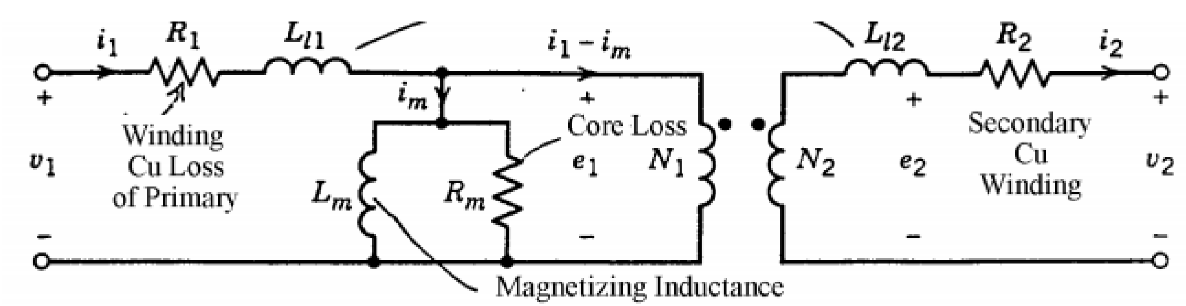
For IL < I(sat) we get lots of core loss and copper loss occurring before core saturation occurs and L= 0.

Over the past twenty years the core loss due to hysteresis and eddy currents in cores has improved primarily by the use of new core materials. Low loss cores are now available up to 5 Mhz.



For Im<< Isat we don't have a catastrophe, but we still create core loss via □Im swings in the core. We model this combination of hysteresis and eddy current loss by an equivalent Rm in parallel with Lm.

The equivalent transformer circuit model then has two currents ILm for reactive current in the core and IRm which models all core losses



3.2.5. The creation of magnetizing current Im

$$i_{primary}(\text{from a load}) \approx \frac{N_2 i_2(load)}{N_1}$$

The magnetizing inductance draws current from the input voltage also, so the total current flowing is:

$$i1 = iprimary + i_m$$

$$i_1 - \frac{N_2 i_2}{N_1} \approx i_m$$

Despite i1 and i2 are both large, Im can still be small, until zero.

The magnetizing current causes a corresponding core flux

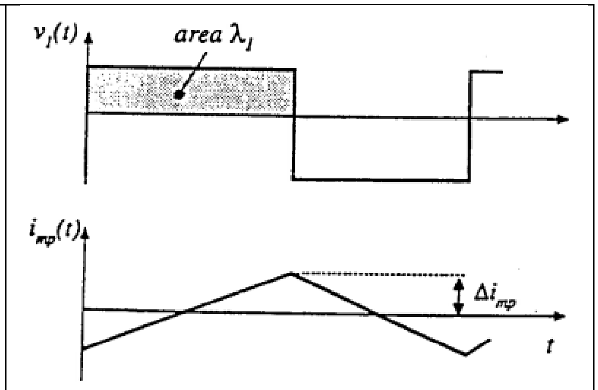
$$\text{which is } \phi = \frac{[N_1 i_1 - N_2 i_2]}{R_c} = \frac{N_i i_m}{R_c}.$$

in a well designed transformer the V1(across the primary winding) waveform alone drives Im. If we assume R1 and Ll1 are small so that the full input voltage appears across LM. Then we can say:

$$i_m = \frac{\int V_1 dt \;\}\lambda_1}{L_m}$$

$$\text{Using } L_m\, i_m = N_1 \phi_c$$

$$B_c = \frac{\lambda_1 = \int V_1 dt}{Z\ N_1\ A_c}$$



We are always conscious that B(core) should never exceed B(saturation), which constrains the absolute number of copper turns possible on the primary.

To reduce Bc the number of wires turns in the primary can be increased. This does not effect N2/N1 provided N2 is changed proportionally to maintain the desired turns ratio. As N1, then Bc ~ V1/dt decreases. In summary, neglecting the core winding area constraint, to keep B(core) < B(sat)

3 .2. 6. Heating Limits of Transformer

At the core surface or inside the center of the wire windings, transformer losses are limited by a maximum 'hot spot' temperature. As we have shown in the first approximation, the temperature rise (° C) is equal to the total power loss (Watts) of the core thermal resistance (° C / Watt).

P total = P(core) + P(windings)

ΔT = R C x PT (core plus copper)

the appropriate core size for the application is the smallest core that will handle the required power with losses that are acceptable in terms of transformer temperature rise or power supply efficiency. We usually cannot exceed a core temperature of 100°. Thermal radiation and convection both allow heat to escape from the core

R θ(total) = R(conv) R(rad)/R(conv) R(rad)= Parallel Combination

Typically we find in practice for cores the thermal resistance varies over
a range: 1 < RQ (total) < 10°C/ W

RQ(total) depends both on core size/shape and thermal constants.

T(core) - T(ambient) = RT (W/°C) PT (total power loss)

Practically, T(core), is never to exceed 100oC since core □(T) and wire insulation degrades.

T(ambient) = 40oC, RT = 10 C/W, and PT = Typically 2-20W
T(core) = RTPT + 40oC = 90 C

3.3 Transformer Model and Applications

3kW-10kW Planar Transformers | Size 560

DESCRIPTION

3kW-10kW Planar Transformers Size 560 with 400A max current rating, frequency range 40-125kHZ. Patented (U.S. Patent 7,460,002) terminals offer mechanical strength and very low resistance. Full Bridge, Half Bridge, Full Bridge ZVS and Push-Pull topologies are included. High efficiency (low loss), low-profile, ultra compact. Outstanding (Pb-free or Pb / Sn Solder) solderability. Standard sizes / customer settings, frequently without start-up or tooling costs, quick custom turn-around. Large secondary pins on terminals reduce the rise in temperature. There are various terminal options (SMD, Thru-hole, screw terminals) available.

SPECIFICATIONS

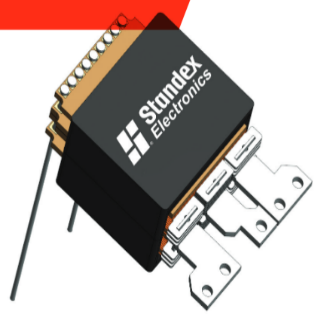
Power Range	**3kW-10kW**
Current Rating Max.	**400A**
Frequency Range	**40-125kHZ**
Isolation voltage pr i-sec/pri-core	**500-5,000VDC**
Topology	**Full Bridge, Half Bridge, Full Bridge ZVS, Push-Pull**

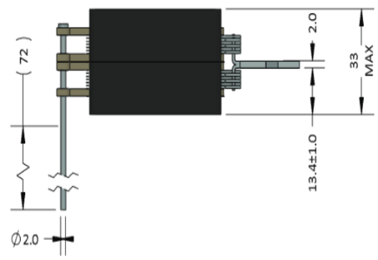
Dimensions L x W x H	**66.0-71.1 x 64.0 x 25.4-30.5mm**
Notes	Length (L) may vary depending on terminals. Height (H) may vary depending on input / output requirements

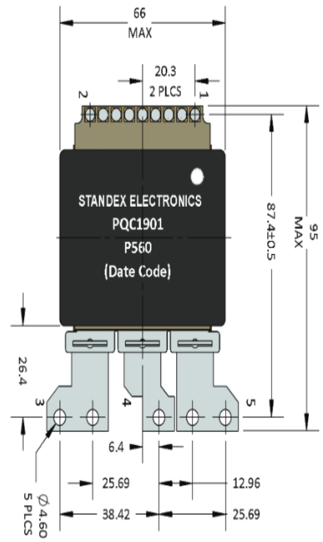
APPLICATIONS

- AC-DC resonant designs
- Aerospace & Military (high reliability/repeatability)
- Automotive
- Electric and Hybrid Vehicles
- Battery Charging (12V, 24V, 48V, 1-10 KW)
- DC-DC Converters (100W-1200W) in distributed power systems
- Distributed Isolated Power
- Feedback Control
- High Current POL Converters
- High Power LED Lighting
- Industrial Power
- Welding
- Isolated Inverters
- Isolated (non-regulated) Bus Converter (Vout 9-12V)
- Renewable Energy – Wind & Photovoltaic Power System, Server – Data Centers (400VDC)

SIZE 560
3kW-10kW
DESIGN EXAMPLE







TRANSFORMER DESIGN | EXAMPLE - PQC1901 (U.S. PAT. 7,460,002)

ELECTRICAL SPECIFICATIONS

Topology	Half Bridge ZVS
Input Voltage	800VDC
Output Power (Output Voltage / Current After Rectification)	6144W Max. (24VDC / 256A)
Turns Ratio - Np / Ns	20T / 1T + 1T
Switching Frequency	50kHz
Duty Cycle, Max.	100%
Efficiency At Full Power (Calculated)	99.24% (47W Losses)
Ambient Temp. Max. (Transfer clamped to heatsink)	+85°C

*Heatsink Provided By Customer

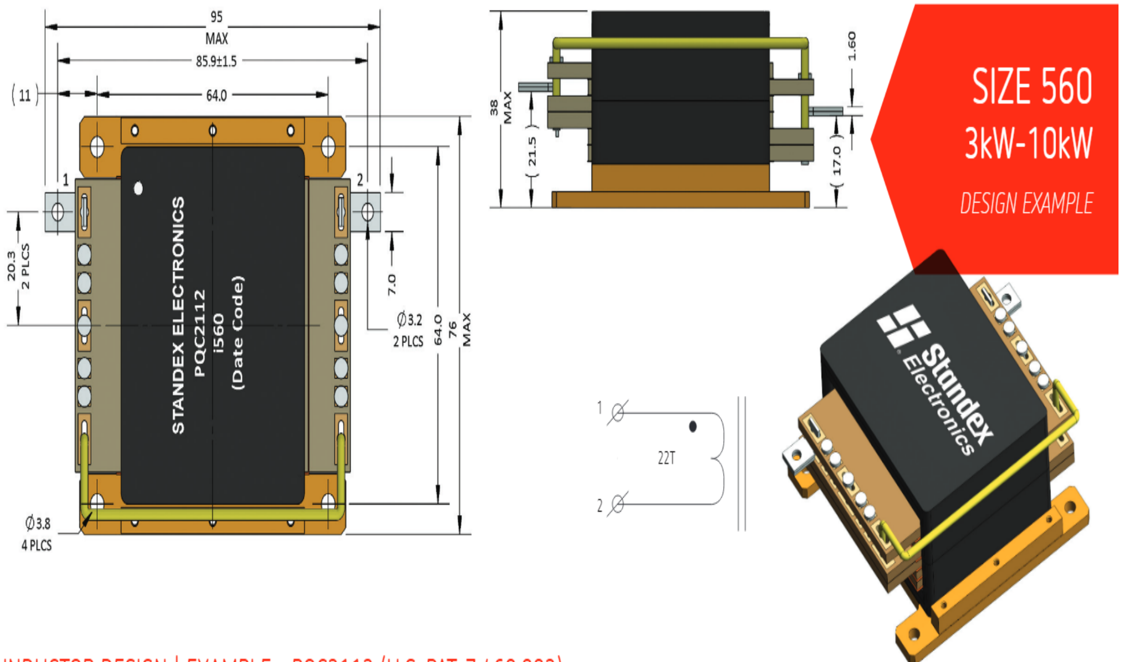
Temperature Rise Hot Spot Heatsink, Max.*	+37.6°C
Minimum Isolation Voltage	
Primary To Secondary And Core	3000VDC
Secondary To Core	500VDC
Primary Inductance, Np, Min.	4000µH
Primary Resistance, Np, Max.	30mOhm
Secondary Resistance, Ns, Max.	0.25mOhm
Leakage Inductance 1-2 / 3-4-5 Shorted, Typ.	3µH
Weight Range	650-700grams

NOTES:

1) FOR OPTIMAL PERFORMANCE A THERMALLY CONDUCTIVE SUBSTRATE BETWEEN FERRITE AND HEATSINK SHOULD BE UTILIZED

2) PATENTED TERMINALS AVAILABLE FOR SPLITTING HIGH CURRENT WINDING

3) CUSTOM TERMINALS CAN BE DESIGNED AND OPTIMIZED



INDUCTOR DESIGN | EXAMPLE - PQC2112 (U.S. PAT. 7,460,002)

ELECTRICAL SPECIFICATIONS

Inductance At Rated Current	100µH ±10%	Temp. Rise Hot Spot Baseplate, Max.	+46°C
Rated Current (Ave. ±12.5A Ripple)	32ADC +3App	Heatsink Temperature Max.	+55°C
Ripple Frequency	100kHz	Resistance Max.	22mOhm
Minimum Isolation Voltage (Winding To Core)	2500VDC	Total Losses At Max. Current	28.7W

NOTES:

1) FOR OPTIMAL PERFORMANCE A THERMALLY CONDUCTIVE SUBSTRATE BETWEEN FERRITE AND HEATSINK SHOULD BE UTILIZED

2) PATENTED TERMINALS AVAILABLE FOR SPLITTING HIGH CURRENT WINDING

3) CUSTOM TERMINALS CAN BE DESIGNED AND OPTIMIZED

4. Resonant Converters with Galvanic insulation

4.1 Galvanic insulations

Galvanic isolation in energy equipment refers to the fact that the output power circuit is electrically and physically disconnected from the input power circuit. Electric isolation is accomplished by using an isolation transformer. Physical insulation ensures that the output power wiring does not contact or connect to the input wiring. Galvanic isolation separates input and output supplies to a system so that energy flows via a field rather than through electrical connections. It enables power to be passed between two circuits that must not be connected.
All personal computers now have galvanic isolation between the input power and the device logic built in. This is a necessity of international protection organizations in order to minimize shock hazards. Therefore, the inclusion of another transformer is redundant. Many people erroneously assume the noise is corrected by galvanic insulation on ground (earth) cables. This is not reliable. This is not accurate. In fact, all galvanic isolation transformers isolate only the power wires, but pass through the ground wire directly.

Galvanic isolation is provided by some UPS systems. Many online UPS units do not have galvanic insulation, despite the common belief that they do. For example, galvanic isolation is not given by on-line models from Exide, Unison, and ON-LINE (Pheonixtec). The standby ON series by Oneac offers galvanic isolation. Therefore, isolation is not a UPS type attribute, but rather a characteristic that can be applied to any UPS. Isolation transformers convert AC current inside a UPS system. Linked devices thus isolating them from the source of power. This protects and suppresses harmful electrical noise against electrical shock.

The real advantage of building an isolation transformer is that there is a substantial reduction in typical mode noise fed to the device. Popular mode noise can also be minimized using noise filters, such as in the APC Smart-UPS series. The filters and the isolation transformer can also work, especially when computers and networks are operating at high frequencies. The isolation transformer is better at very low (audio) frequencies.

Computers and computer peripheral devices are not affected by any audio frequency noise on the power line.

Thus, for computer applications, the isolation transformer has little benefit over filters. The downside of the isolation transformer is the extra heat that shortens UPS battery life if the batteries are in close proximity. Another downside is that the weight of the UPS with an isolation transformer increases significantly.

Typically, transformer-based uninterruptible power supplies are the choice for industrial environments and medical applications that tend to demand Galvanic isolation. Two isolation transformers may, however, be added to the input supplies of a transformer-less UPS, which helps ensure full neutral separation. For a UPS with a single power supply and input, the illustration below describes the simplified installation.

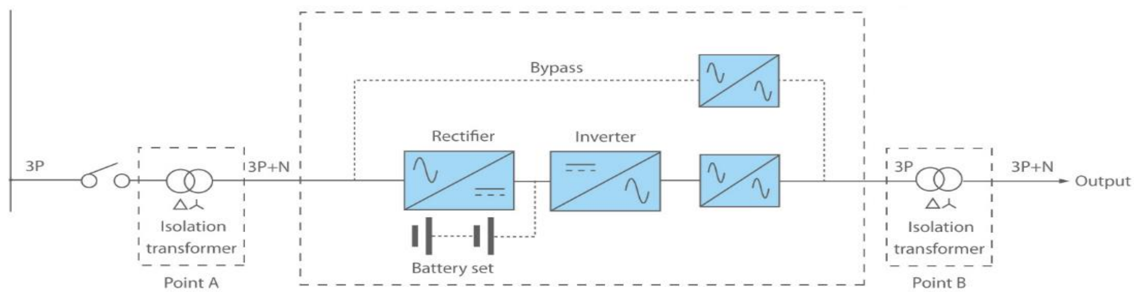


Fig 4.1 Galvanic isolation options for a UPS with a common supply, single input

4.2 Types of Galvanic isolations

Galvanic isolation is the method of separating various sections of electrical circuits to stop current flow while allowing the required amount of data to flow through these sections. In cases where two or more circuits having different soil should communicate with each other, galvanic isolation is used. In the following three types of Galvanic Isolation methods will be discussed.

Inductive or Magnetic Isolation

Digital modulation was used to transform the circuit signals to pulses at the isolation boundary by inductive or magnetic isolation.

Advantages

1. Higher data rate.
2. Smaller size.

Disadvantages

1. Power consumption increases with data transfer rate.
2. Sensitive to the electromagnetic field

Capacitive Isolation

High data rates at lower power consumption.

Advantages

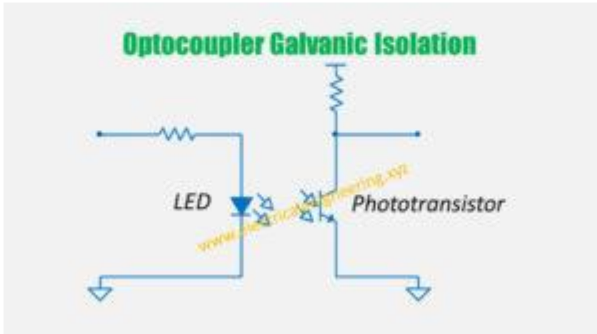
1. Not sensitive to fields
2. High Speed
3. Lower Power Consumption
4. Compact size

Disadvantages

Degradation of lower frequency performance

Optocoupler Galvanic Isolation

The most common component for isolating circuits is the optocoupler or opto-isolator. An LED and a phototransistor are essentially included in the optocoupler. The phototransistor detects the light produced by the LED.



Advantages

1. Good for low-speed communications
2. Works well with digital circuits
3. Cost effective method

Disadvantages

1. Low-speed device
2. Nonlinear for most analog measurements
3. Slow response time
4. Not recommended for high-speed applications
5. Performance change with the lifetime

4.3 Advantages of Galvanic Isolation

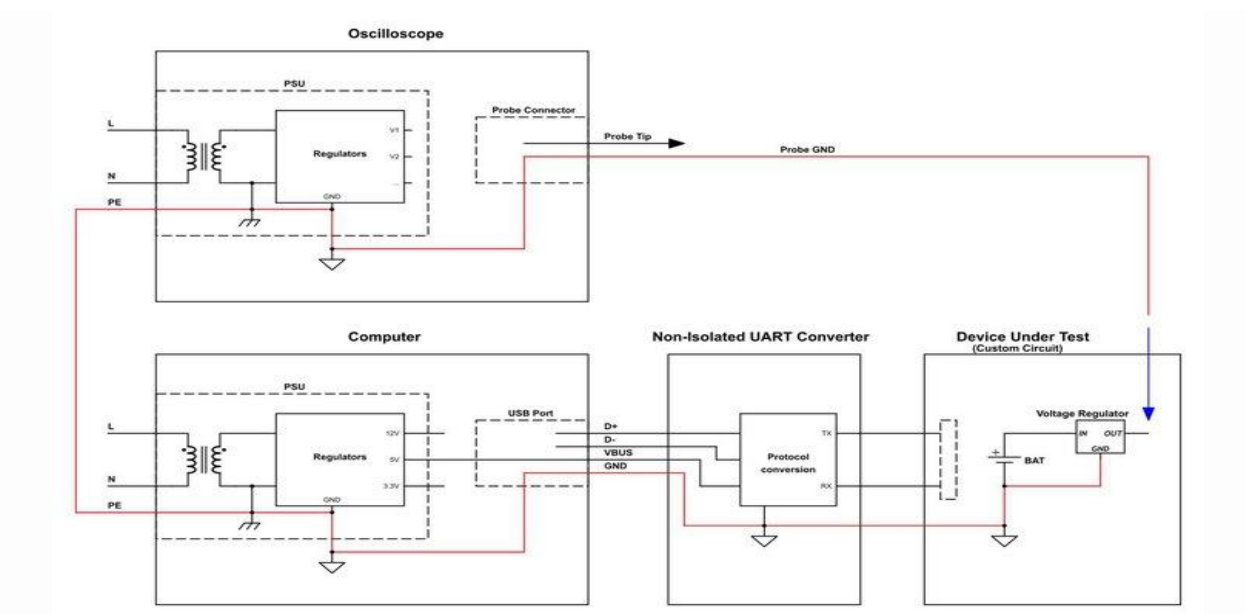
Floating reference

The isolated portion of the circuit is "floating" with respect to the power source. There is no way to keep it consistent with any relation relative to the power source, because there is no clear path for the current to flow back to the power source.

A battery-powered instrument such as a multimeter is a typical case of galvanic isolation, as it has nothing to repair its reference potential, it will float to any reference it is linked to; for example, a battery-powered multimeter may be connected directly to the mains for measurements. its" 0V "is at the potential of the mains, and its battery is" mains potential + 9V. Nothing changed locally on the multimeter.

Safe hardware probing

Without galvanic isolation, the ground (GND) of a circuit is electrically connected to the GND lead of the USB port through a low impedance path. What many people don’t realize, is that a computer’s USB port is often mains earth referenced, meaning the USB GND is actually connected to earth of 230 V power grid. And this is still true even though your computer has an isolated power supply. This is perfectly normal, safe, and is actually how it should be.



5. Application in charging infrastructures for EVs

In this chapter I will discuss resonant converter application related to charging methods for EVS

5.1 Introduction

In terms of high efficiency, less switching losses, LLC resonant converters have many advantages when compared to other converters. It is also capable of operating on a narrow switching frequency where it is possible to achieve zero current switching. This converter is designed for the 15V-20V output range with a 30V input, with an efficiency ranging from 88% to 92%. The two significant attributes of the resonant converters are low conduction loss and low switching loss. In the application of battery charging, this characteristic makes the proposed topology. The battery used in hybrid electric vehicles should be in a way that it would satisfy specifications as smooth and fast charging, high efficiency and high-power density.

The most commonly used battery charging architecture is shown in fig. 5.1 It consist of mainly two stages, namely power factor correction PFC stage and DC-DC converter stage. The power factor correction stage is a continuous conduction mode of boost topology.

The DC-DC converter, which plays an important role in the battery charger by controlling the output current and voltage, is the main focus of this application. The battery's characteristics depend on this stage.

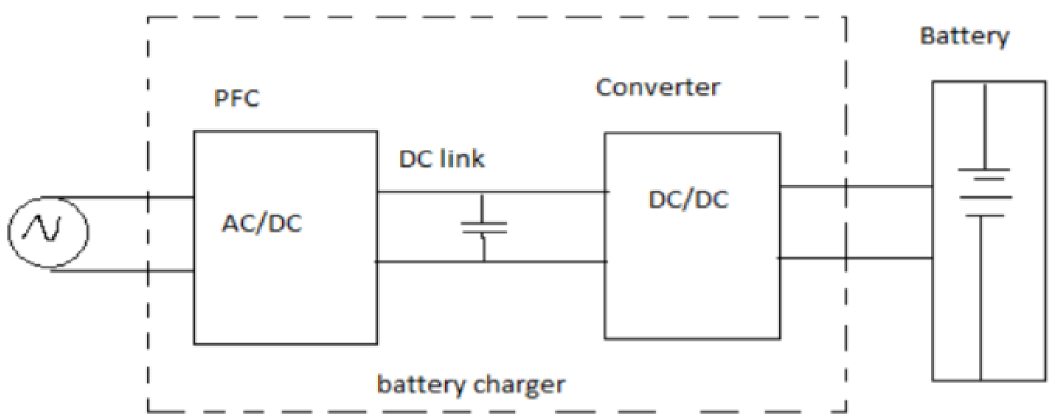


Fig 5.1. Block Diagram of Battery Charger.

The resonant converters can generally be in the form of LLC or LCC configurations. The LLC network where two inductors and capacitors are present, is represented in Figure 5.2.a. The LCC network is represented in Figure 5.2.b, where two capacitors and an inductor are used. In comparison with the LCC network, LLC networks are able to achieve soft switching over a wide range of operations. By combining inductors as transformers, the size of the converters can be decreased in the LLC network.

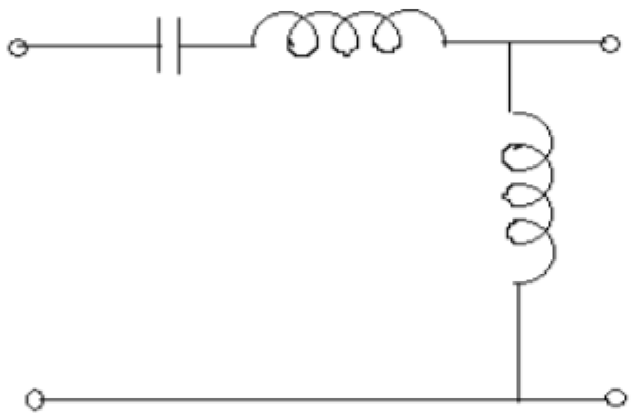


Fig 5.2.a LLC resonant tank

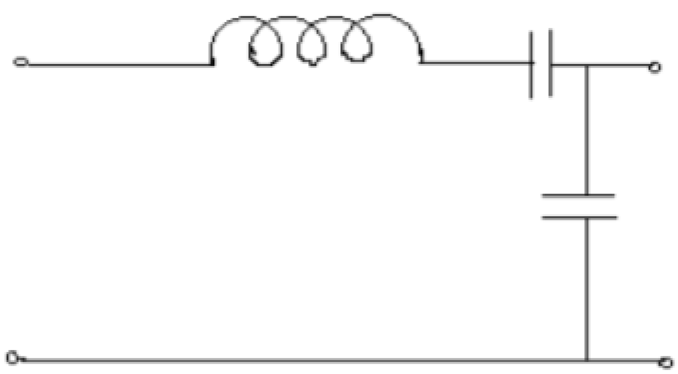


Fig 5.3.a LCC resonant tank

The preferable converter for that application would be the resonant LLC converter and for that application the half series resonant LLC converter is proposed shown as in Fig 5.4

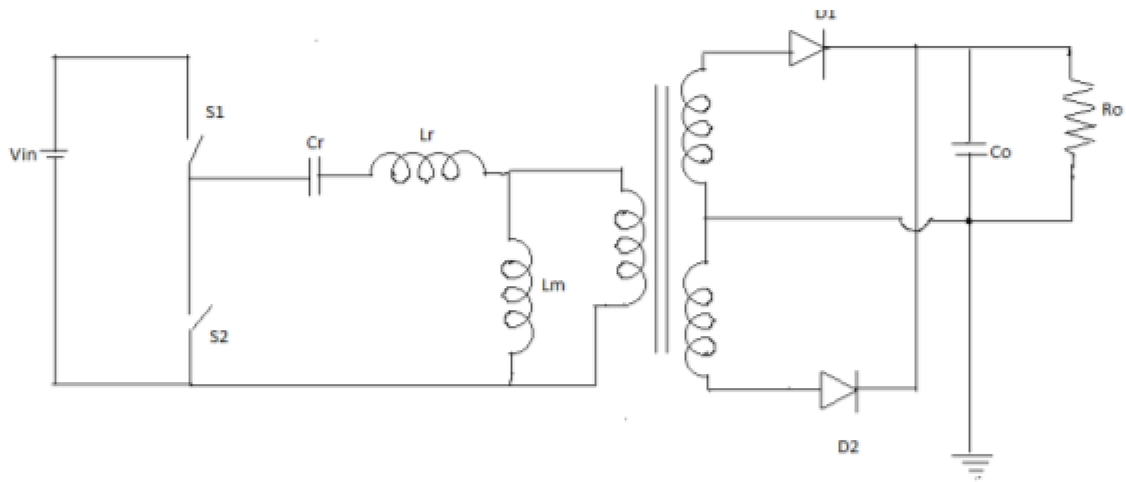


Fig 5.4 proposed resonant LLC resonant converter

5.2 The Operation of LLC series resonant converter

The energy supplied to the resonant tank load is regulated by varying the switching frequency and by controlling the resonant tank's impedance. Two frequencies run on the LLC converter. The first includes Lr, Cr and the second is Lr, Cr, Lm

$$fr1 = \frac{1}{2\pi\sqrt{LrCr}}$$

$$fr2 = \frac{1}{2\sqrt{LrCrLm}}$$

The first equation is the frequency with no load mode, and the second equation is the frequency with load conditions. LLC converters are designed so that fr1>fr2. The frequency depends upon the transformer 's gain. The power flow from the input to the load side is controlled by the frequency of switching. The LLC converter has three modes of operation, depending on the frequency range:

I. At resonance
II. Below resonance
III. Above resonance

- **At Resonance**: The converter is said to operate in resonance condition when the switching frequency is equal to the resonant frequency. The resonant current will be equal to the magnetizing current during this period when the S1 switch is turned off, where there will be no power transfer between the source and the load. Due to the dead time switching between the switches, zero current switching is obtained.

- **Below Resonance**: If the switching frequency is higher than the resonant frequency, the converter will be in the resonant mode below. Here, before the switch switches off and the power is supplied to the load, the resonant current will be equal to the magnetizing current. If the fr1 frequency is less than the switching frequency, it is possible to acquire zero voltage switching throughout the operation, if fr2 is greater than the switching frequency, due to high switching losses, zero voltage switching may be lost.
- **Above Resonance**: The converter is in the above resonance mode if the switching frequency is lower than the resonant frequency. The conduction losses will be reduced due to the increase in switching frequency as the circuit operates in the mode of continuous conduction. As the fr1 is larger than the switching frequency, it is easy to obtain the ZVS even in

5.3 Circuit Operation

For one switching cycle the resonant converter operates in four different modes. For each half cycle two modes of operation takes place.

Mode 1: The voltage across the switch should be equal to zero prior to turning the MOSFET switch on. The resonant current starts to increase across the inductor Lr when the switch S1 is ON. The magnetizing inductance that is coupled to the side of the load will also be charged, but resonance is not involved. The secondary side diode D1 will therefore be ON and energy is delivered to the load. This mode finishes when the resonant current becomes equal to the magnetizing current.

Mode2: The resonant current will be equal to the magnetizing current when S1 is ON in mode 2. The magnetizing inductor will be linked to the resonant circuits in series in this mode, where the output load will be disconnected from the input. Therefore, no power will be delivered to the output, and only the primary side of the transformer will circulate current. This mode ends when the S1 switch is turned off. Mode 1 and mode 2 are repeated as mode 3 and mode 4 for the next cycle, when the MOSFET switch S2 is in ON condition.

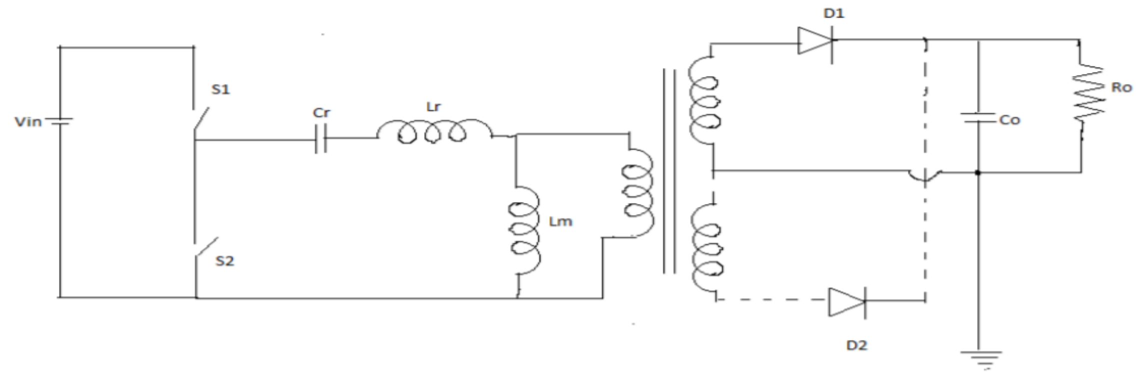


Fig 5.5 S1 ON-mode 1

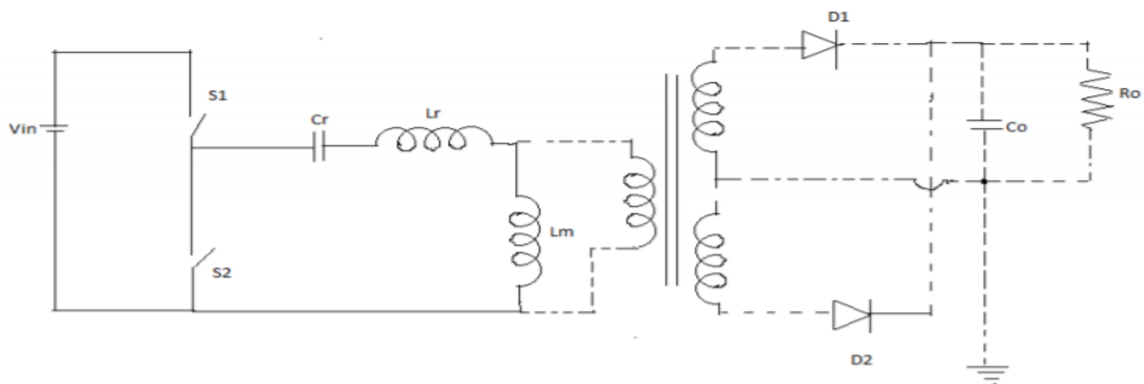


Fig 5.6 S2 ON-mode 2

5.4 DESIGN PROCEDURE

1st: System specifications:

The efficiency of the system should be calculated to determine the input power. If the efficiency data are not given then we can assume the efficiency between range 0.88-0.92 for low voltage and 0.92-0.96 for high voltage and the power input will be:

$$Pin = \frac{Po}{Eff}$$

2nd: we need to determine the maximum and minimum voltage of the converter. The gain value can be set to be between 2-5. Now the minimum resonant voltage is:

$$Mmin = \frac{k}{k+1}$$

$$Mmax = \frac{Vin(\max\ )}{Vin(\min\ )} * Mmin$$

3rd : we need to find the turns ratio of the transformer as the full wave rectifier is connected with the secondary side of the transformer. The turns ratio is calculated as

$$n = \frac{Vin}{(Vo+Vf)^2} \quad \ldots\ldots\; Vf = 0.6$$

4th: Load resistance Calculation:

The load resistance is calculated by: $R_{ac} = \frac{8\, n^2 v_o{}^2 Eff}{n^2 v_o}$

Where, n is the transformer turns ratio
V_o is the output voltage

Eff is the efficiency

5th. Calculation of resonant parameters

$$\mathrm{C_r} = \frac{1}{2\pi \mathrm{Q} \mathrm{f_o} \mathrm{R_{ac}}}$$

$$\mathrm{L_r} = \frac{1}{4\pi^2 \mathrm{f_o}^2 \mathrm{C_r}}$$

$$\mathrm{L_m} = \mathrm{k}\mathrm{L_r}$$

5.5 Simulation*

S. NO	PARAMETERS	VALUES
1	V_{in}	30V
2	V_{out}	15-20V
3	f_s	20kHz
4	f_o	25kHz
5	L_r	190. 09uH
6	C_r	0. 21uF
7	Gain(k)	0. 5
8	Mmin	0. 33
9	Mmax	0. 528

Table 5.1 Input values for simulation

The simulation is done using PSIM software and a constant output voltage is obtained using the PI controller.

The output voltage is obtained as 13V.

The main aim of this topology is to obtain soft switching to decrease the switching losses which in Fig 5.6 is shown as the current through the switch is zero and the voltage starts to increase across the switch where soft switching is obtained.

From fig 5.7 it is deducted that the current flowing through the inductor is sinusoidal because of the capacitor Cr. If the current through the inductor is sinusoidal the voltage across the capacitor should be the same.

in Fig 5.8 it is shown the primary voltage of the transformer.

The proposed converter is a DC-DC converter. So, both input and output should be in a constant DC voltage.

Figure 5.9 shows the 13V constant output voltage of the converter and a current of approximately 1.09 Amps. The controller's reference value is compared to the output from the system's open loop, and error signals are sent to the PI controller to adjust the value of the error. The signal pulses are generated with the help of PWM and a constant voltage of about 16 V is obtained, as shown in Figure 5.11. The output voltage of the converter is effective in the closed loop system as compared to the open loop system in Figure 5.9 and Figure 5.10.

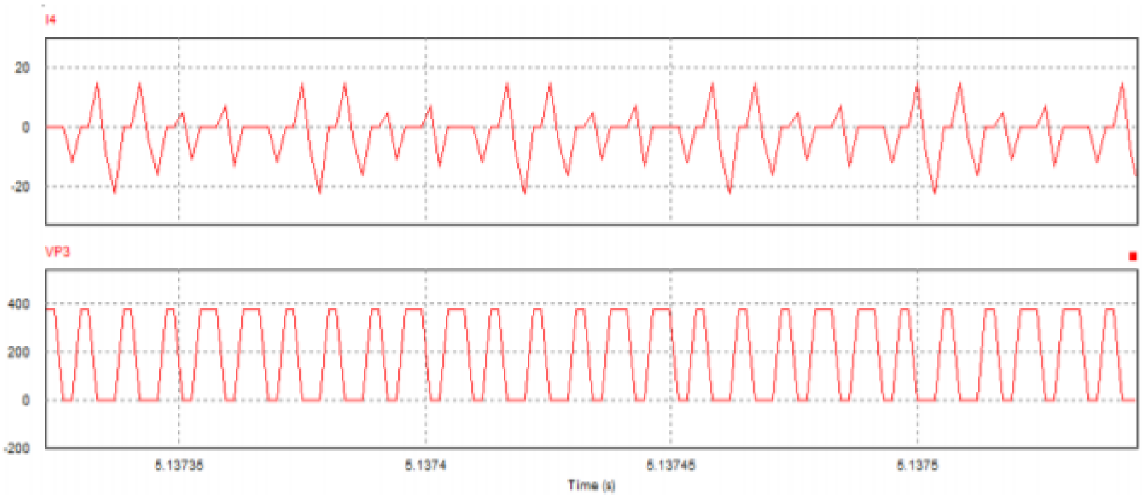


Fig 5.7 Soft switching (ZCS) of the converter

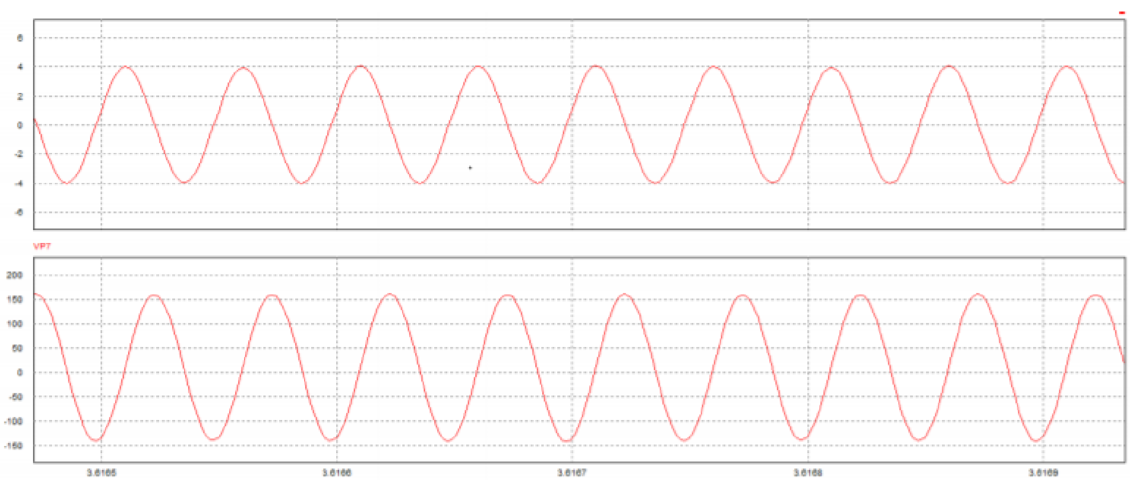


Fig 5.8. voltage across capacitor Cr and current through inductor Lr.

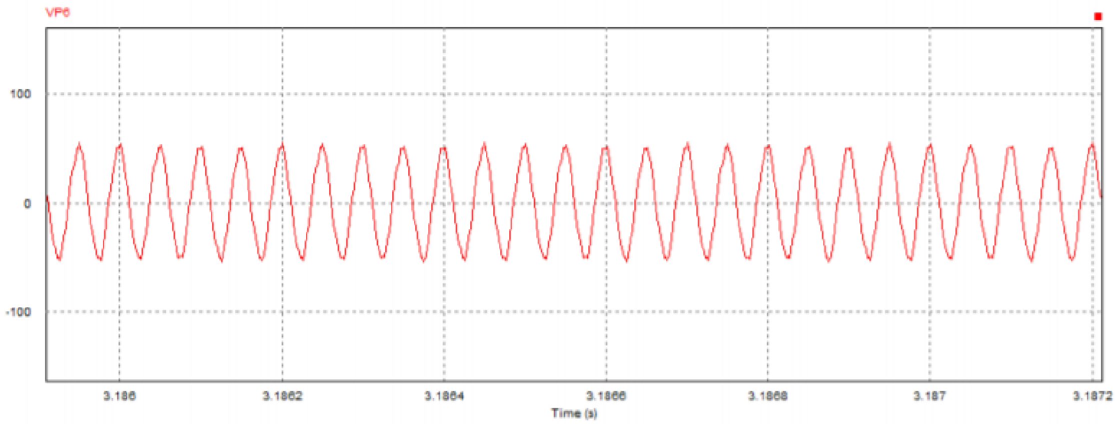


Fig 5.9 Voltage across the primary side of the transformer

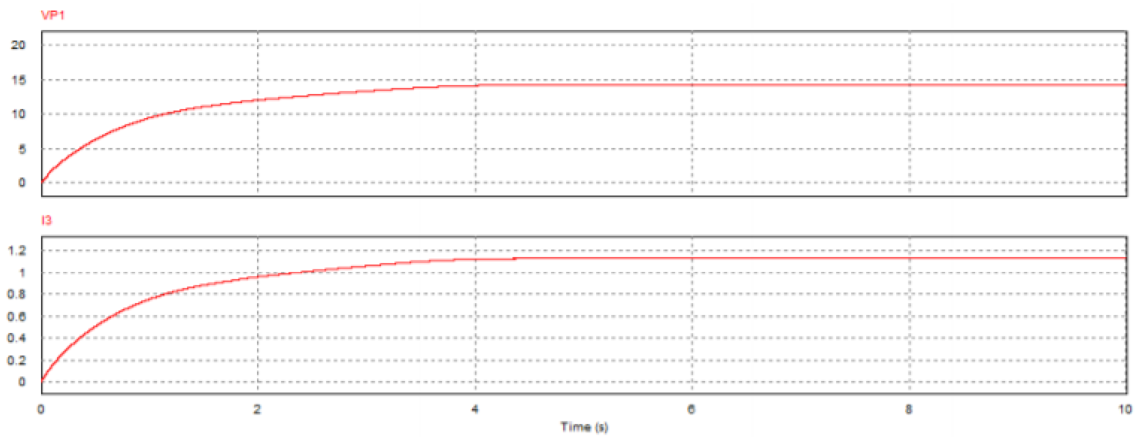


Fig 5.10 Voltage and current waveforms for the open loop system.

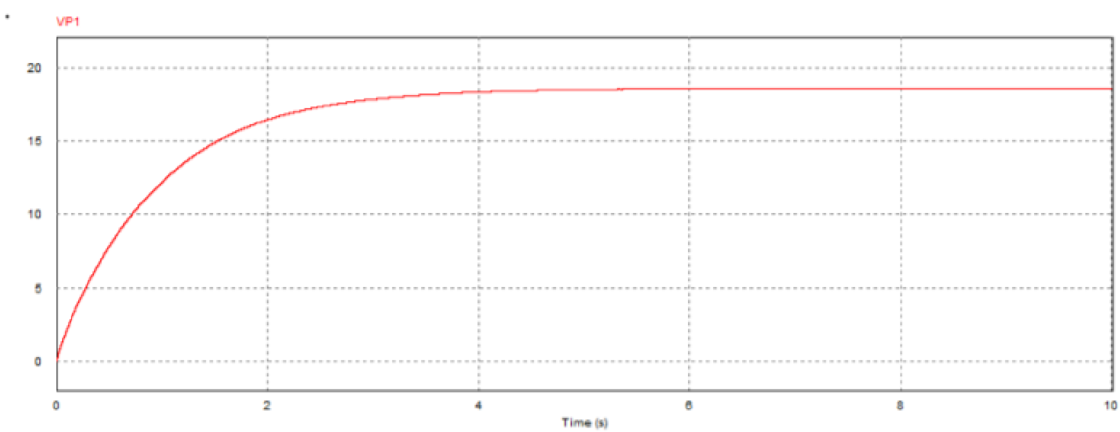


Fig. 5.11 output voltage for closed loop system

*The simulation was deducted from (international Journal of Electrical Engineering. ISSN 0974-2158 Volume 8, Number 4 (2015), pp. 379-388)

Conclusions:

LLC resonant converters have many advantages when compared with other converters in terms of high efficiency, less switching losses. It is also capable of operating in narrow switching frequency where zero current switching can be achieved.

This paper reviewed the resonant converter with insulating transformers and wide range of voltage discussing the main features of the resonant converters and its topologies, comparing the different operating modes ZVS and ZVC, as well as reviewing applications on soft switching and Applications using resonant converters. Similarly was done with High Frequency Transformers, explaining their basics, topology, and showing some models. Moreover, in the next chapter Galvanic isolations were discussed on how these isolations operate, various types of the isolations were shown, same with the advantages and disadvantages of Galvanic Isolations.

In the last chapter, a review on an application of resonant converters in battery charging for Electric Vehicles. For that application ZCS based LLC resonant converter is used, PSIM software was used for simulations and the results are discussed. From the waveforms it is inferred that a constant output voltage is obtained without any distortion because of the filter capacitor used across the load. Therefore, resonant converters can be used for battery applications because of their constant output voltage with reduced switching losses.

Bibliography:

1. DESIGN CONSIDERATIONS FOR HIGH FREQUENCY TRANSFORMERS Agile Magnetics
2. https://standexelectronics.com/products/3kw-10kw-planar-transformers-size-p560/
3. Resonant Circuits and Soft Switching (LLC Resonant Converter and Resonant Inverter) Toshiba Electronic Devices & Storage Corporation
4. https://www.coilmaster.com.tw/en/a1-9418/High-Frequency-Transformer.html#:~:text=A%20transformer%20is%20an%20electrical,impinging%20on%20the%20secondary%20winding/
5. Power Electronics and Motor Drives Systems Book 2017
6. Survey of Resonant Converter Topologies Reproduced from 2018 Texas Instruments Power Supply Design Seminar SEM2300, Topic 1 TI Literature Number: SLUP376
7. https://www.pantechsolutions.net/blog/introduction-to-resonant-converter/ Pantech Solutions – Introduction to resonant converters
8. https://www.se.com/ww/en/faqs/FA157465/ Description of Galvanic Isolation (Schneider Electric)
9. Advantages, Limitations and Costs of Galvanic Isolation https://warmcat.com/2016/10/07/advantages-limitations-and-costs-of-galvanic-isolation.html
10. https://www.electricalengineering.xyz/power-electronics/3-types-of-galvanic-isolators/
11. Georgin-Galvanic insulation http://www.georgin.com/pdffiles/fc-guide-interfaces-en.pdf
12. "Flyback Converter," [http://www.infineon.com/cgi/ecrm.dll/ecrm/scripts/sol_ov.jsp?oid=16941]. April 2003.
13. Mohan, N., Underland, T., and Robbins, W., Power Electronics, Converters, Applications, and Design, John Wiley & Sons, Inc., New York, NY, 1995.
14. Sangtae, I., "Forward Converter Type PC SMPS with FPS," Technical Report, Fairchild Electronics Co., LTD, Korea. 2002.
15. "Topologies," Technical Report, Power One®. 2000.

16. "Half-Bridge Push-Pull Converter," [http://henry.fbe.fhdarmstadt.de/smps_e/hgw_hilfe_e.html]. May 2003.
17. R. Beiranvand, B. Rashidian, M. R. Zolghadri, and S. M. H. Alavi, 'A design procedure for optimizing the LLC resonant converter as a wide output range voltage source, IEEE Trans. Power Electron, vol 27, no., 8, pp. 3749-3763, Aug 2012.
18. H. Bai, Y. Zhang, C. Semanson, C. Luo, and C. C. Mi, "Modeling, design and optimization of a battery charger for plug-in hybrid electric vehicles, IET Elect. Syst. Transp., vol. 1, no. 1, pp. 3-10, Mar. 2011.
19. F. Xiang, H. Haibing, Z. J. Shen, and I. Batarseh, "Operation mode analysis and peak gain approximation of the LLC resonant converter, IEEE Trans. Power Electron., vol. 27, no. 4, pp. 1985-1995, Apr. 2012.
20. H. Bai and C. Mi, "Comparison and evaluation of different DC/DC topologies for plug-in hybrid electric vehicles, Int. J. Power Electron., vol. 4, no. 2, pp. 119-133, Feb 2012.