Reverse-conducting IGBTs for induction cooking and resonant applications

About this document

Scope and purpose

The aim of this document is to introduce the working principle of induction cooking and to provide insights on the IGBT technologies from Infineon that are best suited for induction cooking applications. In addition, it provides basic guidelines for the design of the target application. The document extends the scope also to inverterized microwave oven applications.

Intended audience

Engineers who need to design resonant converters and select IGBTs for induction cooking applications.

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1 Introduction

The principle of induction heating was discovered by Michael Faraday in 1831. In an experiment with two coils wired around an iron core, he discovered that during the switching event of a battery connected to the first coil an opposite current flow could be measured with a galvanometer on the second coil. He concluded that an electric current could be produced by a changing magnetic field. Since there is no galvanic connection from the primary to the secondary coil he called the voltage in the second coil “inducted.” His law, established on the basis of this experiment, is called a general law which states that the “EMF electro-motive force inducted in a circuit is directly proportional to the time rate of change of magnetic flux though the circuit.” Heinrich Lenz formulated it later in this way: “The polarity of the inducted EMF is such that it tends to produce a current that will create a magnetic flux to oppose the change in magnetic flux through the loop.” [1]

This principle was used for transforming different levels of voltage for the efficient transmission of electricity and the operation of electrical machines. An effect observed during the energy transformation was the heat generated in the cores, which are generally made of laminated stacks of steel. In many practical applications, the heat generation is an undesired phenomenon, and therefore all electrical machines are designed to minimize it. However, there are systems whose main goal is to produce heat (e.g. in industrial furnaces). For these systems, the electromagnetic field can be used for induction and heat generation without direct thermal contact. Such an application is therefore called induction heating and, in case it is applied to cooking purposes, it is normally referred as inductive cooking.

a) b)

Figure 1 Typical appearance of induction cookers: a) single- and b) multi-hob.

The amount of heat that can be generated by means of the induction heating phenomenon depends on several factors like the frequency of the magnetic field, the material used in the heating vessel, the coupling between the magnetic field and the vessel. Generally, the higher the frequency and amplitude of the magnetic field driven in the vessel, the more heat is produced in the vessel itself. In state-of-the-art designs, a range of frequency between 20 kHz and 75 kHz is sufficient to guarantee heating power up to 4 kW for most ferromagnetic materials. In order to extend the range of operation to non-ferromagnetic materials (e.g. cast iron or even aluminium), higher frequency and higher magnetic field strength are usually needed, thus increasing the frequency range up to 100 kHz. Such a magnetic field is generated by means of a power inverter, which stimulates a coil with a current oscillating at the required frequency. IGBTs are the most common power semiconductor devices used for controlling the current in this frequency range, due to their high current capability. Due to the limitation of the power semiconductors, switching losses and the resulting thermal stress on the components, today’s household cookers utilize, for the highest power rating, operation frequencies of 20 kHz to 40 kHz. The main reason for moving towards higher switching frequencies is to reduce the audible noise generated by the coils and to allow for a wide operation of different materials. An alternative solution using power MOSFETs based on silicon or wide-bandgap materials could achieve higher operating frequencies but are still less attractive because of higher cost.

In Figure 1 several commercial example of induction cookers are shown. According to the U.S. Department of Energy, the 25% savings in primary energy can increase to 60% in process heating in today's induction heating appliances [1]. Improving the degree of efficiency of inverters in commercial and residential applications requires advanced control systems and new generations of power semiconductors. To optimize the electrical behavior of these devices, new generations of semiconductors are being continuously developed by Infineon Technologies. IGBTs from 650 V to 1600 V are widely used for induction heating appliances. The reason for such a high range of voltage classes is mainly related to different circuit topologies that are used to build the high-frequency inverters. In the following the most common topologies are presented, as well as the main features of Infineon Technologies IGBTs targeting each of them.

2 Induction heating principle

As already mentioned in the previous section, the alternating magnetic field, which is needed to induce the heat generation in the vessel, is generated by means of a current circulation in a physically exciting coil [2]. Flat coils, sometimes referred to as “pancake coils” are used for induction cookers with flat surfaces. Nevertheless, the coil can be shaped in different manners in other types of appliances. Typical coil arrangements are shown in Figure 2. In most cases, the coils utilize ferrite bars to increase the confinement and a better alignment of the magnetic field. The design and the optimization of a coil is generally performed with the aid of 3D CAD and electromagnetic simulation software [3].

a) Typical “pancake coil” for countertop cooking appliances

b) Typical “bowl-shaped” coil for rice cooking appliances

Figure 2 Different coil shapes for induction cooking appliances.

For any given coil, the relation that links the magnetic field and the stimulating current is the following:

$$\Phi = L * I,$$

where Φ is the magnetic flux across the coil surface, and L the equivalent inductance of the coil. From the previous equation it is evident that the amount of magnetic flux generated from the coil is proportional to the amplitude of the current that flows across it. As long as there is no other conductor placed in the sourrounding area of the coil, the magnetic field does not produce any energy transfer, and the energy that is absorbed by the coil is mainly reactive. If a vessel is placed on top of the coil, the magnetic field starts to penetrate to its bottom surface, as shown in Figure 3a. In this case, the magnetic field that is coupled with the vessel has two effects:

- It generates in the vessel an electro-motive force (EMF) proportional to the time rate of change of the magnetic field itself, according to the Faraday-Lenz law:

  $$E = -\frac{d\Phi_c}{dt},$$

  where Φ_c is the magnetic flux that is coupled with the vessel. As a consequence of the conductive material, the applied EMF produces a current[1] that heats up the bottom surface of the vessel.

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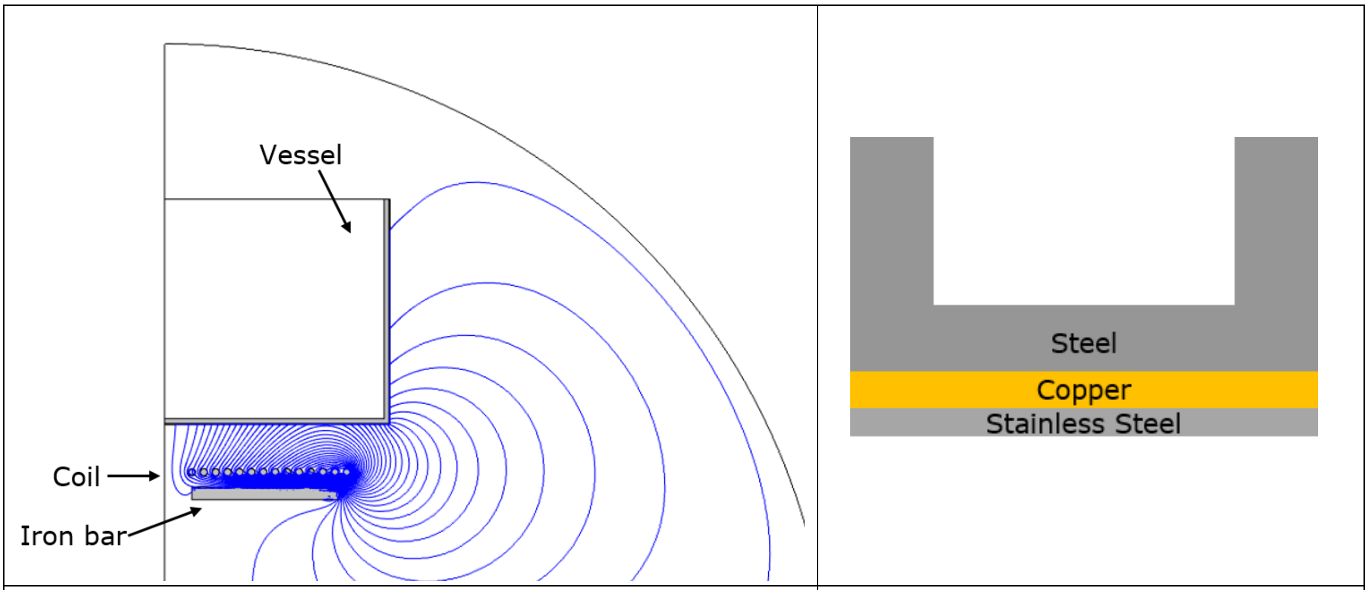
[1] This current flows in closed circles that are perpendicular to the magnetic field and therefore is usually called “eddy currents.”

- If the vessel is made of material with strong magnetic sensitivity, e.g. any ferromagnetic material, the oscillating magnetic field produces losses because of magnetic hysteresis. The heat generation in this case is proportional to the area of the hysteresis curve of the specific material.

An example of a simulated magnetic field distribution is shown in Figure 3a for a vessel with a material structure depicted on the right[1]. As can be seen, the highest field density is generated in the region between the coil and the vessel. The figure also illustrates the effect of the bars made of magnetic material, which have the benefit of shielding the electronic components, usually placed below the coil, from the magnetic field. The bars have also the additional benefit of confining the magnetic filed to a smaller area, thus increasing even more the field density inside the bottom of the vessel.

As a consequence of the previous analysis, it is evident that an increase in heat generation can be achieved by either an increase of the amplitude or frequency of the alternating current, or with the increase of the magnetic coupling between the magnetic field and the vessel. The contribution of the two effects described above add up to generate the heat dissipation in the bottom of the vessel. Since typical cooking vessels are made of very good conductive materials, the main contribution to the heat generation comes from hysteresis. For this reason, to heat up typical non-ferromagnetic materials like aluminium, a magnetic field is required with a much higher amplitude and frequency compared to ferromagnetic materials, given the same coupling with the vessel.



a) b)

Figure 3 **a) Simulated magnetic field lines with a coil and vessel arrangement; b) Schematic representation of the vessel materials.**

As mentioned in the introduction, the coil is characterized by its inductance value, and can therefore be schematically represented by a simple inductor. On the other hand, the vessel can also be considered as an inductor, whose equivalent number of coils is equal to one. With this simplification, the equivalent model of the coil and the vessel arrangement can be represented by means of a simple transformer, whose coupling factor indicates the amount of coupling between the magnetic field, which is generated by the coil, and the vessel.

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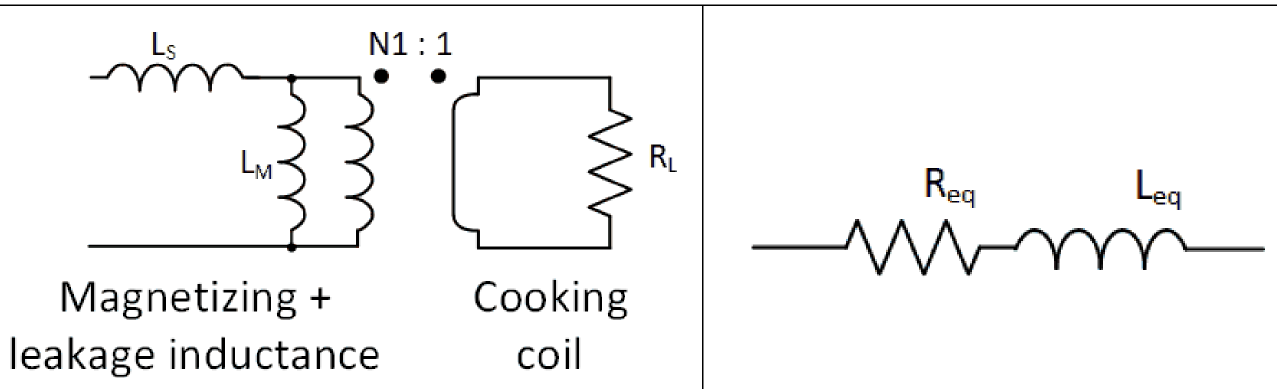
[1] The details of the simulation geometry and the boundary conditions are voluntarily omitted in this overview, as the main purpose of the author is to provide only a reference to the magnetic coupling between the coil and the vessel. More detailed analyses, which would require better knowledge of the actual information about the coil design and the vessels, is beyond the scope of this document.



a) Transformer model (neglecting parasitic resistance and core losses)

b) Series load model

Figure 4 **Equivalent model of the coil and vessel arrangement.**

As for a real transformer, in order to improve the coupling factor, it is necessary to reduce the distance between the coil and the vessel as much as possible. In typical induction cookers, the vertical distance between the coil and the glass surface is in the range of 1-3 cm, and the most important contribution to magnetic coupling is given by the alignment between the vessel and center of the coil. As will be shown later, it is always beneficial to have a non-unity magnetic coupling, as the resulting leakage inductance can be used in combination with a capacitor to produce the oscillating current by means of so-called resonant inverter topologies. These topologies are presented in the next section. In Figure 4, two different models of the coil and cooking vessel arrangement are presented, namely the loosely coupled transformer method, which is physically more accurated but more complicated to use, and the series load model, which is much easier to use and is derived from the first model by neglecting the effect from magnetizing inductance. The equivalent model of the coil and vessel can be obtained by means of simulation or by measuring directly the impedance at the coil terminals. In both cases the variation of the parameters with the temperature must be explicitly considered.

3 Resonant converter topologies for induction cooking applications

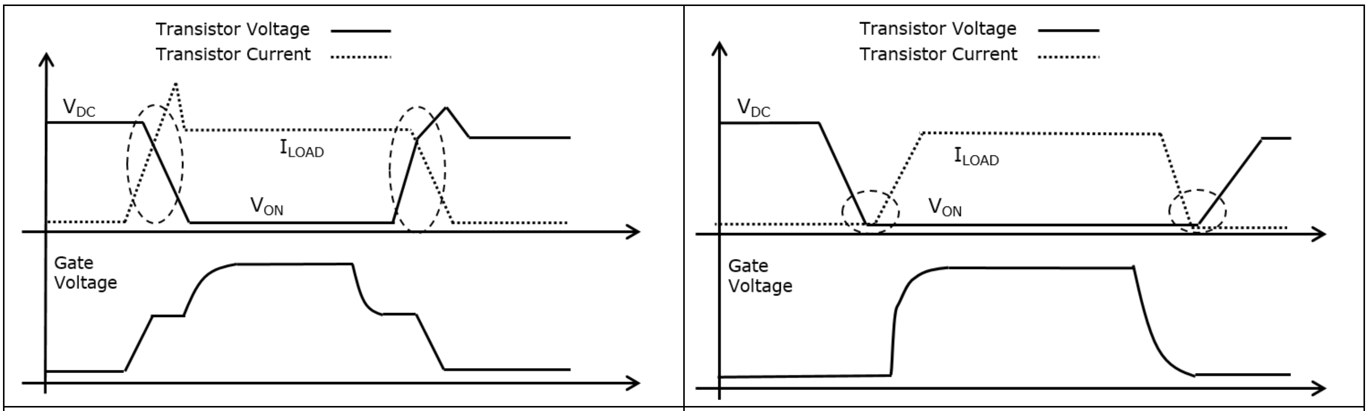
There are mainly three different topologies used to build an induction-based cooking appliance [4]. They include:

- Quasi-resonant converter, also know as single-ended parallel resonant converter: mainly used in single-hob stoves and rice cookers, which are especially popular in Asia. It is also used in the low-end segment of multi-hob cookers and in inverterized microwave ovens.
- Half-bridge series resonant converter: mostly used to build high-end multi-hob cookers and microwave ovens.
- Full-bridge series resonant converter: usually limited to the commercial cooker market, due to the higher power capability.

In the following, the analysis is limited to the first two topologies, as they are the ones mainly used in domestic induction cooking applications. Before analyzing in detail the operating principle of each topology, it is worth dedicating a few words of introduction to the concept of hard- and soft-switching operations.

3.1 Hard-switching and soft-switching operation of power semiconductor devices

Modern power electronics converters are built with one or more power switches that continuosly commutate from conduction to off-state, and vice versa. In most applications there is a certain time interval during each commutation where the current and the voltage of the switch are both non-zero, as shown in Figure 5a, which leads to a significant power dissipation associated with each commutation. This operation is usually referred as hard-switching commutation mode. As a consequence, the swithing frequency of the power converter is usally limited by the amout of losses produced during each commutation. In general, increasing the switching frequency is beneficial for reducing the size of the passive components. In order to increase the operating frequency beyond the limit posed by the switching losses, a power converter can be designed in such a way that the power switches operate with so-called soft-switching commutations [5]. In this type of commutation, the switches commutate either when its voltage is zero (ZVS, zero-voltage switching) or when its current is zero (ZCS, zero-current switching). In some configurations, both the ZVS and ZCS are possible at the same time. A typical ZVS waveform is shown in Figure 5b.



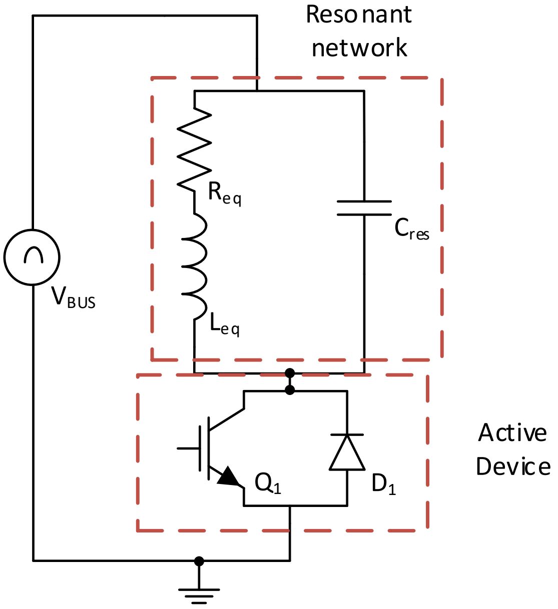
a) b)

Figure 5 **Typical hard-switching (a) vs ZVS (b) waveforms.**

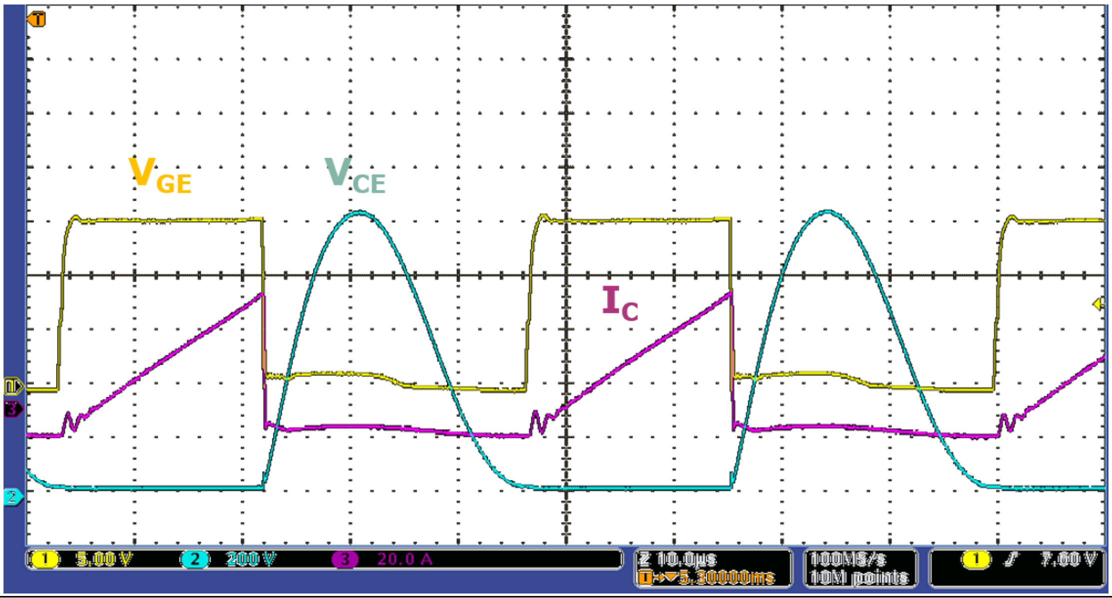
In order to achive soft-switching operations for the power switches, resonant converters are used. These converters are built with resonant tanks to create an oscillatory current and voltage that lead to ZVS and/or ZCS operations. A resonant converter can be efficiently used in induction cooking applications, since the same oscillatory current can be used to produce soft-switching commutations, and at the same time to generate the magnetic field in the vessel for heating purposes.

3.2 Quasi-resonant (QR) converter

The schematic of a typical quasi-resonant converter [6] is shown in Figure 6a. The topology consists of a single IGBT, with an antiparallel diode, which is connected to the resonant tank. The resonant coil is the same type used to produce the magnetic field in an induction cooker. Typical switching waveforms of the IGBT are shown in Figure 6b. When the switch is on, the current flows first in the diode and then it moves to the IGBT as soon as it changes its polarity. During the commutation from diode to IGBT, the current becomes zero and voltage is always very low, therefore a ZVS and ZCS turn-on is achieved. At the IGBT turn-off, the current drops very fast, whilst the collector voltage increases much more slowly, as it is limited by the resonant capacitance. Therefore a virtual ZVS operation can also be achieved at turn-off.



a)



b)

Figure 6 **Schematic of a quasi-resonant converter and typical IGBT waveforms .**

Figure 7 shows the four operating phases of a quasi-resonant converter during a switching cycle [8]:

- Phase I: the first phase starts when the current in the inductor changes polarity, and consequently with the IGBT turn-on. During this phase, the inductor current (and therefore the IGBT current) increase almost linearly with a rate of rise given by the ration between bus voltage (VBUS) and the resonant inductor. The phase I ends when the IGBT turns off.
- Phase II and III: in these phases, the current flows between the inductor and the capacitor. As a consequence, the inductor current and the capacitor voltage resonate with the natural frequency of the resonant tank. Phase II ends when the inductor current becomes negative whilst phase III ends when the voltage on the switch becomes negative, and the diode starts to conduct.
- Phase IV: the diode turns on as soon as the voltage across the switch becomes slightly negative. In this phase, the current in the active device is negative (it flows from the emitter to the collector terminal) and it decreases in absolute value. The rate of change is almost constant and it is approximately equal to the rate of rise in phase I.

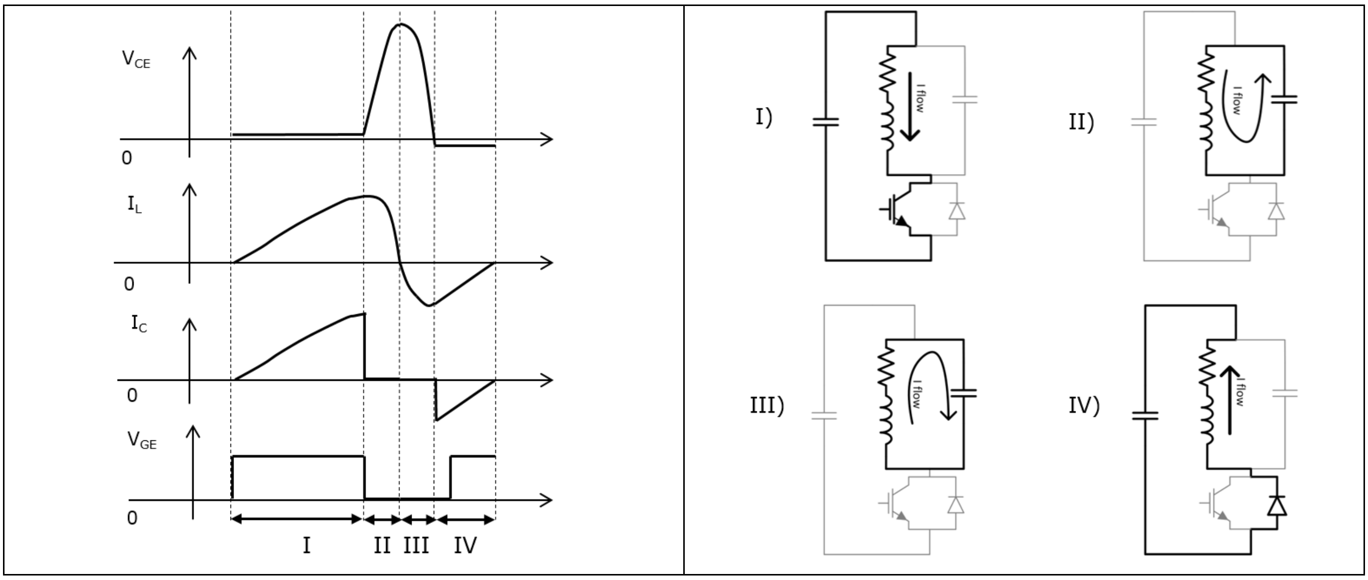


Figure 7 Operating states of the quasi-resonant converter.

The power control in the quasi-resonant converter can be achieved by varying the conduction time of the IGBT. A longer IGBT conduction time results in higher peak current in the resonant inductor and consequently higher power delivered to the vessel.

3.2.1 Basic equations of QR converter

The basic equations that describe the operation of the QR converter for a given specific switching cycle are shown in the following. Parameters *R* and *L* represent the values of the equivalent resistance and inductance of a specific coil and vessel arrangement, whilst the parameter *C* represents the value of the resonant capacitance. Parameter V_{BUS} is the voltage across the bus capacitor, which is assumed to be constant during the given switching cycle.

- In Phase I and IV the circuit behaves as a typical RL series circuit, therefore the current can be described by the following equation:

$$I_C = I_L = I_0 + (I_{MAX} - I_0) * \left(1 - e^{-\frac{t}{\tau}}\right) \tag{1}$$

where $I_{max} = \frac{V_{BUS}}{R}$, $\tau = \frac{L}{R}$ and I_0 is the turn-on current of the diode.

The peak current of the IGBT can be expressed as a function of the IGBT on-time, T_{ON}:

$$I_{IGBT,peak} = I_0 + (I_{MAX} - I_0) * \left(1 - e^{-\frac{T_{ON}}{\tau}}\right) \tag{2}$$

- In Phase II and III the circuit operates as a typical RLC series network with the initial current equal to the turn-off current of the IGBT. Assuming an oscillatory damped behavior [7], the typical equation of the current can be written as:

$$I_L = A * e^{-b} * \sin(c * t + \gamma) \tag{3}$$

where $b = \frac{R}{2*L}$, $c = \sqrt{\omega_0^2 - b^2}$, $\omega_0 = \frac{1}{\sqrt{L*C}}$.

The coefficients A and γ have to be determined according to the boundary conditions:

$I_L(0) = I_{IGBT,peak}, I_L(T_{OFF}) = I_0$

As a first approximation, one can calculate the maximum peak voltage of the IGBT, when in off-state, starting from the maximum current in the inductor, by using the following formula:

$$V_{CE,peak} = \sqrt{\frac{2*E}{C}} \tag{4}$$

where E is the energy stored in the resonant inductor, which is approximately equal to $E = \frac{L*I_{IMAX}^2}{2}$. It is evident from equation 4 that the IGBT voltage at turn-off has a linear dependence on the inductor current and a square root dependence on the inductor value. However, the above formula is derived from the approximation that the equivalent resistance is zero, so the actual peak voltage is indeed smaller.

3.2.2 Hard-switching operation in QR converter

ZCS and ZVS IGBT turn-on operation is achieved in a QR converter as long as the inductor current can flow in the diode. This can only happen if the IGBT voltage becomes negative during the resonant oscillation. In the case of too little energy being stored in the inductor at the IGBT turn-off or if there is a high equivalent resistance of the vessel, the IGBT voltage might not reach zero. In this case, the IGBT can no longer be switched at zero voltage, although it is still at zero current, as shown in Figure 8. During turn-on of the IGBT there is a large current spike due to the fast discharge of the resonant capacitor. During this phase the current of the IGBT is limited by its equivalent on-state resistance and by the stray inductances of the circuit. This hard-switching event increases the overall power dissipation of the IGBT. The power dissipation during the IGBT turn-on can be estimated as the product of the residual energy stored in the resonant capacitance, when the IGBT turns-on, and the switching frequency of the IGBT:

$$P_{diss,turn-on} = 0.5 * C * V_{ON}^2 * f_{sw} \tag{5}$$

where V_{ON} is the residual voltage in the resonant capacitance when the IGBT turns on. The turn-on spike can be limited by the usage of a lower gate voltage, e.g. 10 V[1]. The height of the pulse will be lower resulting in a better EMI behavior of the system. A voltage step from turn-on to the optimal gate voltage should be considered in the system to have an increased current capability of the system.

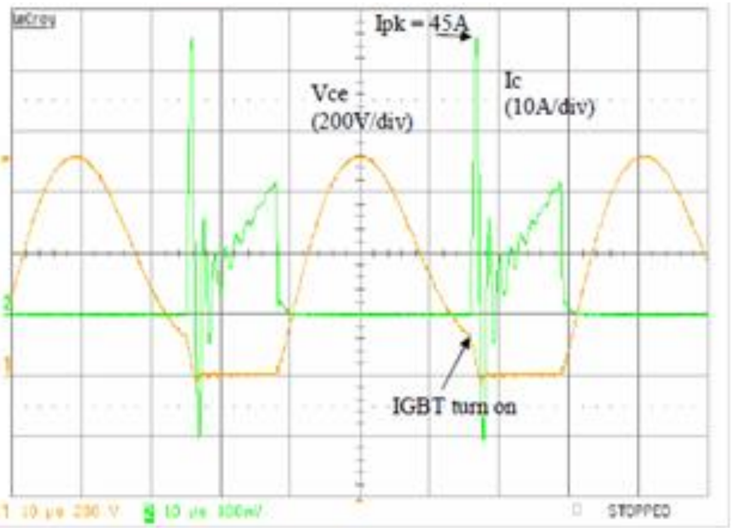


Figure 8 **Hard-switching operation in QR converter.**

When the systems operates in hard-switching mode, the power dissipation of the IGBT increases signifcantly thus also reducing the conversion efficiency. As this condition is predominant at low output power, one possible solution to avoid a hard-switching condition is to adopt a so-called burst mode. In this operation

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[1] Whilst this measures can reduce the amplitude of the IGBT current spike, it doesn't reduce the energy dissipated in the IGBT, since this is only dependent on the residual voltage across the resonant capacitance and on the capacitance value.

regime, the minimum output power of the system is set in such a way that no hard switching operations occurr, and a lower output power can be achieved by an on-off modulation of the inverter, as shown in Figure 9. The example of a possible on-off strategy for different power levels is shown in Table 1, with the assumption that a minimum output power of 1 kW is sufficient to avoid continuous hard-switching operations.

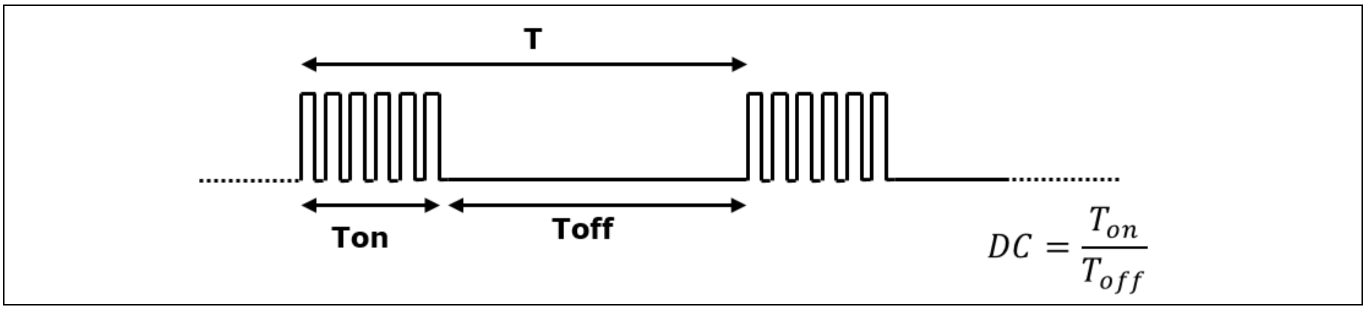


Figure 9 **Burst-mode opearation in QR converter.**

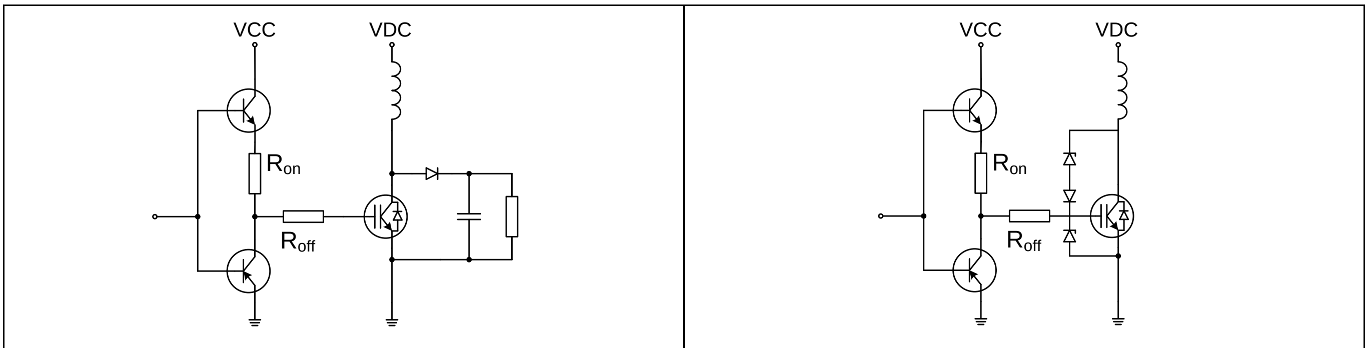
Table 1 **Output power regulation during burst mode**

Desired output power (W)	Inverter on duty cycle
1000	100%
800	80%
600	60%
400	40%

The minimum operation frequency of the burst mode has to be set high enough to avoid any non-uniformities of heat distribution in the vessel, and at the same time to avoid noticeable fluctuactions of the temperature in the food, which would end in unefficient cooking processes. On the other hand, the maximum operating frequency is usually limited by the EMC requirements. Usually, a burst frequency of 0.2-0.3 Hz is used.

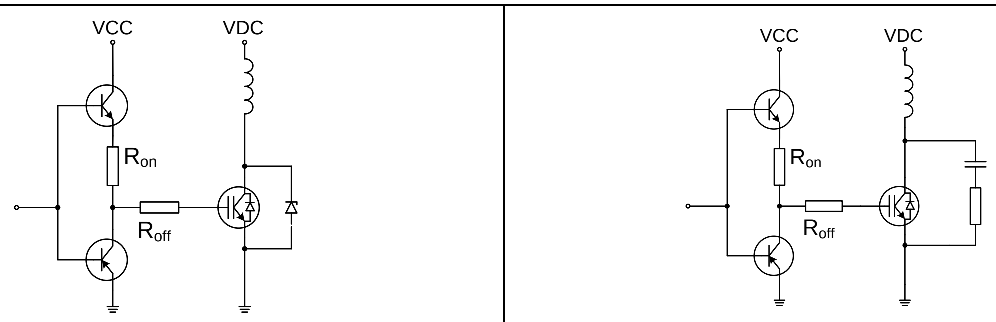
3.2.3 Overvoltage protection

A serious problem in the single-ended topology is that the peak collector voltage of the IGBT is not fixed but depends on the resonant load and on the output power. In addition, unpredictable events on the grid, like surges or mains interruption, can significantly increase the peak voltage. To reduce the voltage surge, a varistor is commonly used, although it can’t always absorb all the additional energy produced by grid fluctuations. For this reason, the IGBT voltage class should be selected to have adequatly safe voltage margin.



Snubber network 1

Active clamping network



Power clamping diode

Snubber network 2

Figure 10 **Possibile solutions to limit the collector-emitter voltage of the IGBT.**

In addition, other protection mechanisms could be considered. As an example, one possibility to limit the overvoltage on the IGBT is to turn it on actively with a series of Zener diodes connected between the collector and the gate terminals (the so-called active clamping mode). Other possibilities to clamp the collector-emitter voltage are shown in Figure 10.

3.2.4 Advantages and disadvantages of the quasi-resonant converter for induction cooking applications

The QR converter is quite popular in the induction cooking market. Some of the benefits for which this topology is chosen include:

- Very good power conversion efficiency due to soft-switching operation of the power switch
- Simple design with only one power switch
- Lower cost compared to alternative converter configurations used in induction cooking

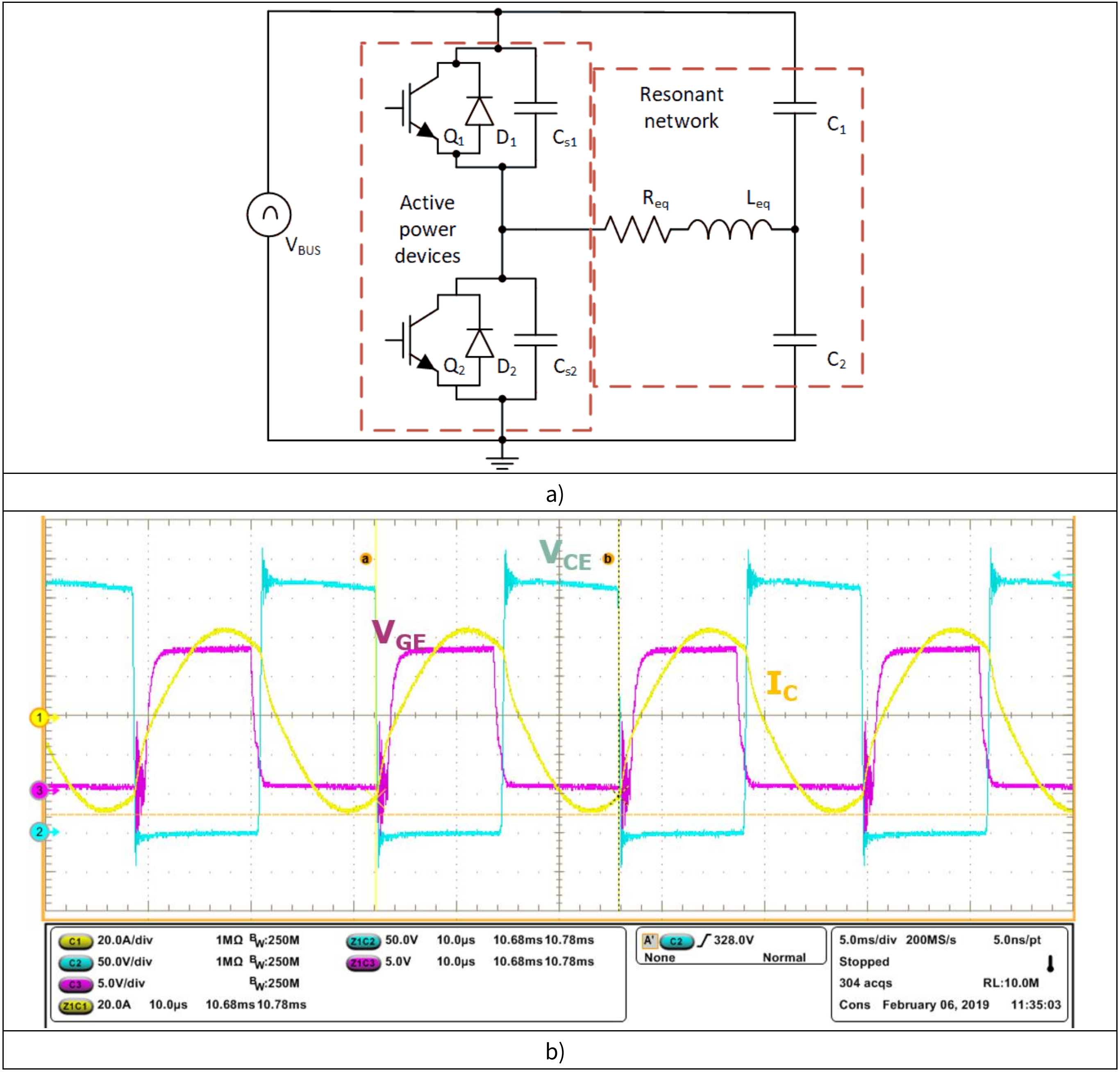
Some of the drawbacks of this topology include:

- Voltage-resonant operation that makes the VCE of the IGBT higher than the input voltage (in the range of a thousand volt)
- High sensitivity to grid-voltage variation (e.g. surges, mains interruption, etc.) as the voltage resonance amplifies the input voltage
- Control difficulties, as the load is unpredictable, which can significantly change the operation of the converter (e.g. damping factor, resonant frequency, etc.)
- Switching frequency cannot be controlled accurately, as it depends on the load
- Maximum power limited by the maximum breakdown voltage of the IGBT
- Minimum continuous power operation limited by the occurrence of hard-switching operation

The QR converter is mostly used in induction cooking appliances that have only one cooking zone. In this case, it offers the best trade-off between cost and performance. Usually the inverter power rating is limited to 2.2 kW, because for larger power, the voltage class of the IGBT would increase, resulting in an increase of device losses. Due to the cost pressure in the consumer market, this topology is becoming more and more popular also in the market for mutli-hob appliances, especially in the low-end segment.

3.3 Half-bridge series-resonant converter

The schematic diagram of a typical half-bridge series-converter [9] is shown in Figure 11a. The main difference with respect to a quasi-resonant converter is that the inverter stage consists of two IGBTs arranged in a totem pole configuration. In this configuration, the maximum voltage on each IGBT can never exceed the voltage of the bus capacitance, regardless of the output power of the inverter. This represents a huge advantage with respect to the QR converter, as low-voltage devices can be used, typically 600-650 V, which have much lower losses compared to high-voltage devices. The typical switching waveforms of an IGBT, operating in this topology, are shown in Figure 11b, where the two devices of the bridge have the same conduction times, and the resonant converter operates in the inductive region[1]. When the switch is on, the current flows first in the diode and then it moves to the IGBT as soon as it changes its polarity. During the commutation from diode to IGBT, the current is zero and the voltage is always very low, therefore a ZVS and ZCS turn-on is reached. When the IGBT turns off, the load current is reverted to the diode of the other device, and the same behavior is repeated in the second semi-cycle.



a)

b)

Figure 11 **Schematic of a half-bridge series-converter and typical IGBT waveforms.**

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[1] That is, when the switching frequency of the inverter is larger than the resonant frequency of the RLC network.

Figure 12 shows the 4 operating phases of a half-bridge converter during a switching cycle [10][8]:

- Phase I: during this phase the high-side (HS) IGBT is on and the current in the load inductor increases. At the end of phase I, HS IGBT is turned off and the load current eventually reverts to low-side (LS) diode.
- Phase I-b: in this phase, the inductor current flows in the snubber capacitances, if present, and the voltage across the LS IGBT is reduced. The phase ends when the LS diode becomes directly biased.
- Phase II: during this phase, the current flows into the LS diode; the current in the inductor is still flowing in the same direction as in phase I.
- Phase III: the current in phase III becomes negative and naturally reverts from the the LS diode to the LS IGBT. LS IGBT turns on and LS diode turns off with zero losses. At the end of this phase, LS-IGBT turns off and the current is diverted into the HS diode.
- Phase III-b: in this phase, the inductor current flows in the snubber capacitances, if present, and the voltage across the HS IGBT is reduced. The phase ends when the HS diode becomes directly biased.
- Phase IV: during this phase, the current flows into the HS diode; the current in the inductor is still flowing in the same direction as in phase III.
- Phase V: at the end of phase IV, the current changes direction again, and the HS IGBT turns on in soft-switching conditions. The HS diode turns off also in soft switching with no reverse recovery effect.

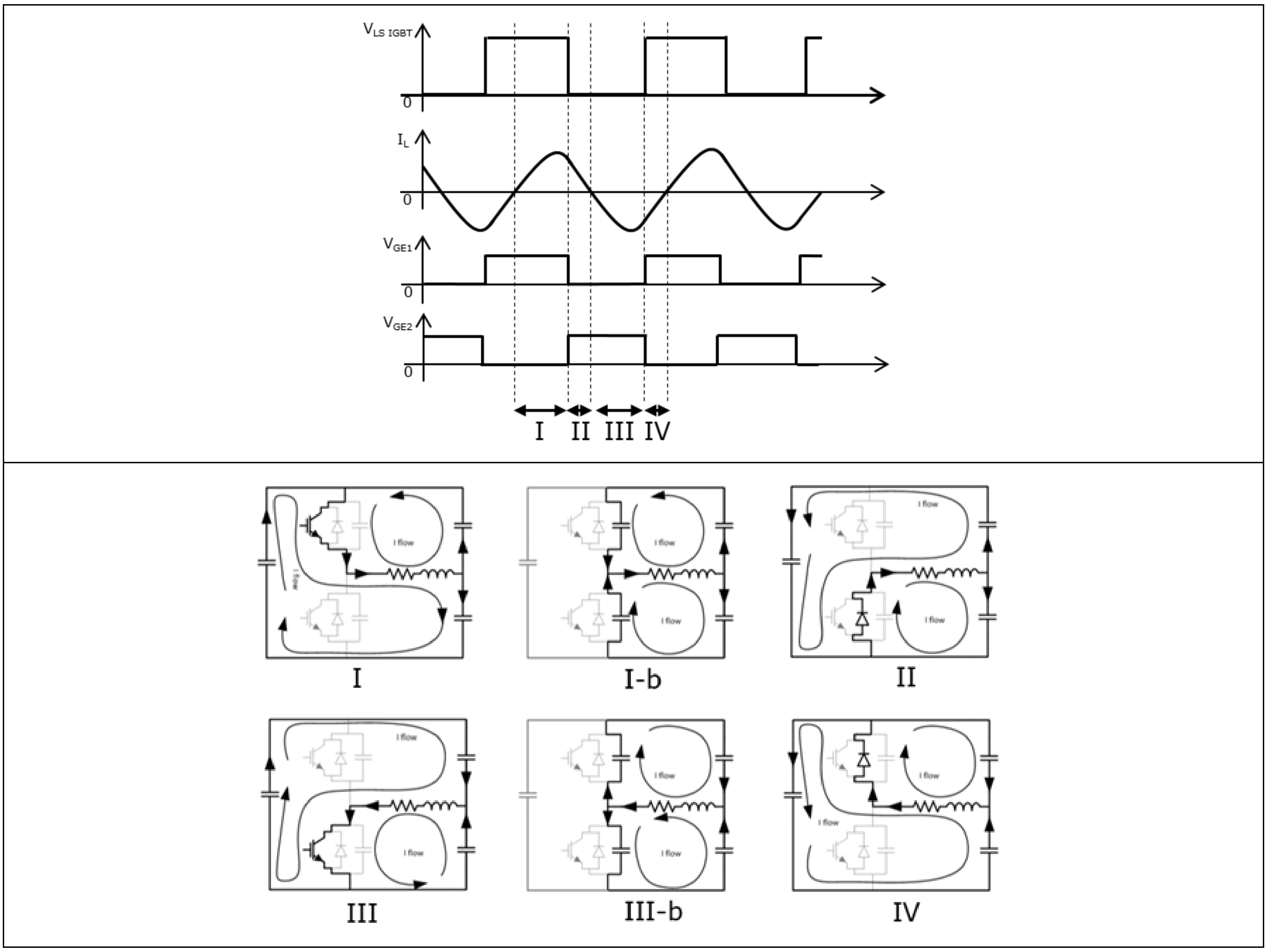

Figure 12 **Operating states of the half-bridge resonant converter. The arrows in the bottom diagram represent the direction of the current for each phase.**

During the power switch turn-off, the rate of rise of the voltage is not limited by the resonant capacitance, as for the QR converter. In this case, the turn-off losses could be relevant if IGBTs with non optimized perfromances are used. However, the switching losses can be reduced by adding a snubber capacitance in parallel to each IGBT. The use of snubber capacitance has mainly two advantages:

- It reduces the switching losses of the IGBT
- It makes the dv/dt more stable and controllable, thus improving the EMI behavior of the switch

A side effect of the snubber capacitance is that it increases the time between the turn-off of one IGBT and the turn-on of the opposite diode. This must be taken into account when designing the proper dead time of the two gate signals[1]. Figure 13 depicts the typical behavior between the turn-off switching losses of an IGBT and the snubber capacitance[2].

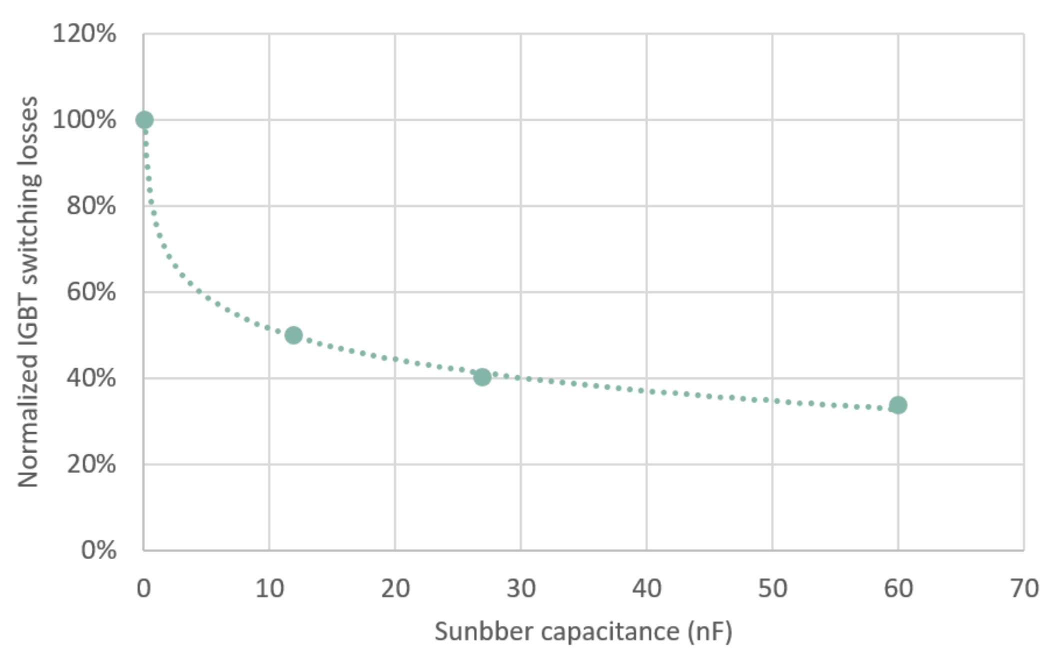

Figure 13 **Typical normalized turn-off switching losses with different snubber capacitance.**

In principle, the half-bridge inverter can operate also in capacitive mode[3]. However, this operation is usally not preferred as it produces higher losses in the IGBT due to the device operating in hard-switching turn-on. During hard-switching turn-on, the losses are usually larger than for turn-off, due to the reverse recovery losses of the diode.

In the half-bridge topology, the output power in the load resistance can be easily controlled via the switching frequency of the inverter. This aspect is important when designing induction cookers with more than one inverter. To avoid acoustic noise, two main aspects have to be considered:

- the switching frequency of each inverter should not be lower than 20 kHz (upper limit of human audible range)

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[1] Alternatively, a better approach, even if more complicated, would be to trigger the IGBt turn-on by the zero crossing of the inductor current.

[2] The reader should note that, whilst the characteristic is considered to be typical for a general IGBT, the actual reduction rate can be very different according to the specific IGBT technology that is considered.

[3] That is, when the switching frequency of the inverter is lower than the resonant frequency of the RLC network.

- if more than one inverter is operating at the same time, the difference in operating frequency must fall outside the audible range

So, as an example, if one inverter operates at 25 kHz, the other inverter has to operate either at the same frequency ±20 Hz or above 45 kHz. In most cases, it is more convenientfor both the inverters to operate at the same frequency. In a half-bridge topology, where the switching frequency can be controlled accurately, it is much easier to achieve a matched frequency compared to the quasi-resonant topology. However, there may be cases where the power requirements of each inverter correspond to switching frequencies that do not fulfill the aforementioned conditions. In this case, several switching patterns (usually protected by intellectual properties) are possible to maintain the iso-frequency operations and to deliver the target power rating to the different inverter, at least as an average.

3.3.1 Basic equations of half-bridge series-resonant converter

The basic equations that describe the operation of the HBSR converter for a given specific switching cycle are shown in the following. Parameters *R* and *L* represent the values of the equivalent resistance and inductance of a specific coil and vessel arrangement, whilst the parameter *C* represents the value of the resonant capacitance. Parameter V_{BUS} is the voltage across the bus capacitor, which is assumed to be constant during the given switching cycle. In order to simplify the analysis, it is worth noting that during a switching cycle, the circuit in Figure 11a is equivalent to the circuit in Figure 14, i.e., a typical RLC series circuit.

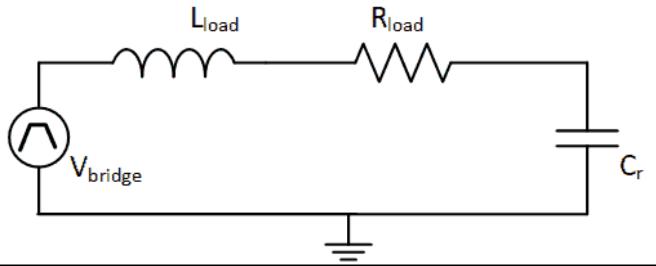


Figure 14 **Equivalent resonant circuit during a switching cycle**

Therefore, the basic equation of the current in the switch is the same as equation 3:

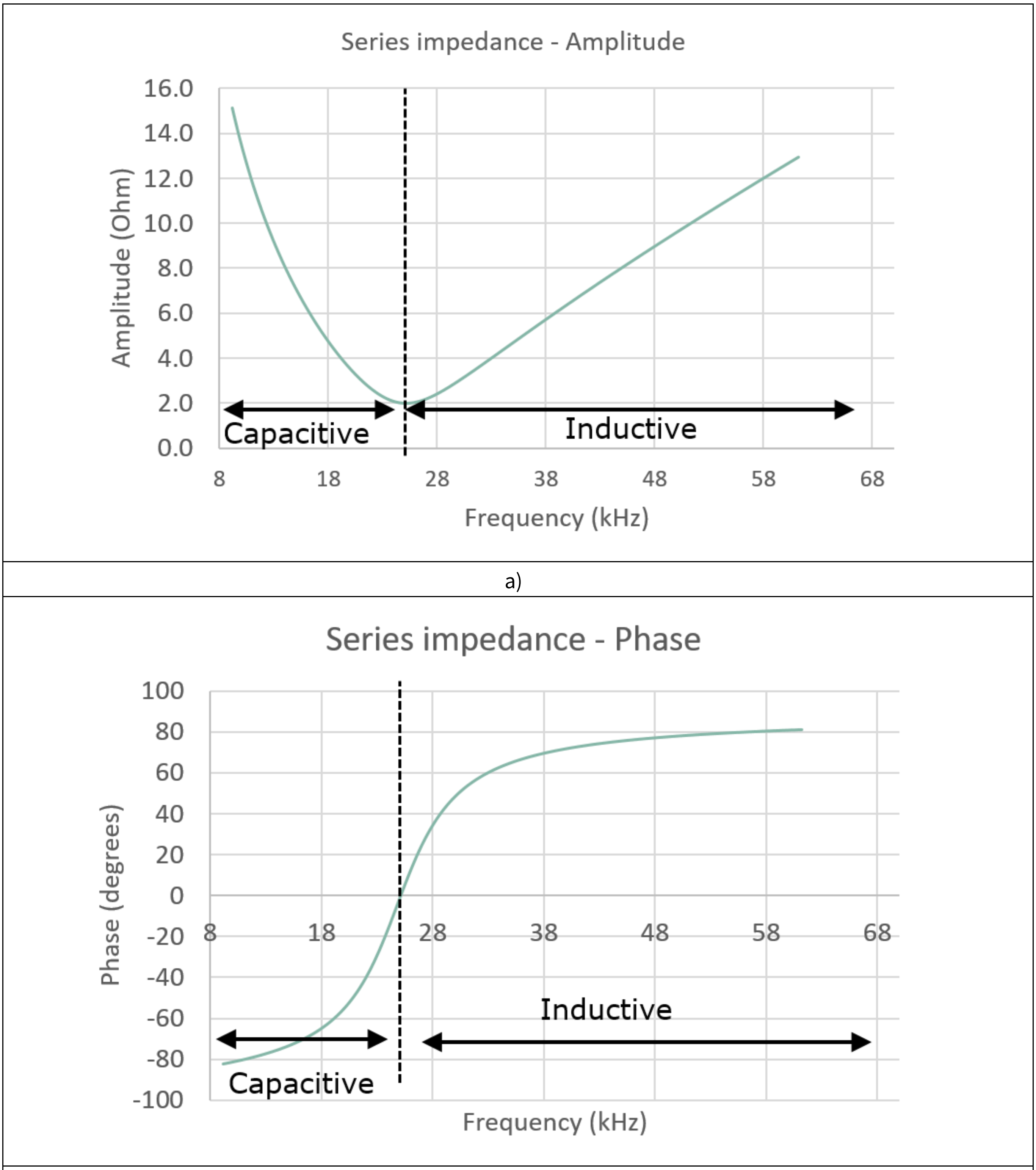
$$I_L = A * e^{-b} * \sin(c * t + \gamma),$$

where $= \frac{R}{2*L},\ c = \sqrt{\omega_0^2 - b^2},\ \omega_0 = \frac{1}{\sqrt{L*C}}.$

The coefficients A and γ have to be determined according to the boundary conditions. In case a duty cycle of 50% is used, the IGBT turn-off current and the diode turn-on current must be the same in absolute value, with opposite signs: $I_L(0) = -I_L(T_{OFF}) = I_{diode,peak}.$

3.3.1.1 Analysis with first-harmonic-approximation method

If a duty-cycle of 50% is used, i.e., the resonant circuit is stimulated by a square wave, an alternative method to analyze the circuit is the so-called first-harmonic-approximation (FHA) also known as sinusoidal approximation [11]. For a relative low damping, the RLC series circuit behaves like a bandpass filter, where the central frequency of the passing band is equal to the resonant frequency. Since the input signal of the network is a square wave, all harmonics that fall outside the passband are significantly attenuated. Therefore, as an initial approximation, one can analyze the circuit considering only the harmonic of the input signal that falls within the passband of the filter. This approximation is more valid the higher the quality factor of the filter, which can be calculated as: $Q = \frac{Z_0}{R} = \frac{\omega_0 L}{R}$. The quality factor is a dimensionless parameter that is expressed as the ratio of the energy stored in the magnetic circuit and the energy dissipated in the damping resistance.



a)

b)

Figure 15 **Amplitude (a) and angular phase (b) of the impedance of the RLC circuit(R = 2 Ω, L = 40 µH and C = 1 µF)**

With this approximation, the amplitude and the phase shift of the current in the resonant tank can be derived by calculating the equivalent impedance of the resonant circuit, which can be written according to the following formula:

$$Z_L = R + j\omega L + \frac{1}{j\omega C} \tag{6}$$

where ω represents the switching pulstation of the input square-wave voltage. The amplitude of the current is therefore calculated as:

$$|I(j\omega)| = \frac{|V(j\omega)|}{|Z_L(j\omega)|} = \frac{|V(j\omega)|}{\sqrt{R^2+(\omega L-\frac{1}{\omega C})^2}} \tag{7}$$

while the phase of the current can be calculated as:

$$\angle I(j\omega) = \mathrm{atan}(\frac{\omega L-\frac{1}{\omega C}}{R}) \tag{8}$$

The values of the amplitude and the phase of the impedance at different frequencies are shown in Figure 15. The current has a maximum at $\omega = \omega_0 = \frac{1}{\sqrt{L*C}}$, which is when the excitation frequency is the same as the natural frequency of the resonant load. The phase shift between voltage and current in this case is equal to zero degrees. The maximum current can be calculated simply as the ratio of the amplitude of the input voltage first harmonic and the value of the damping resistance.

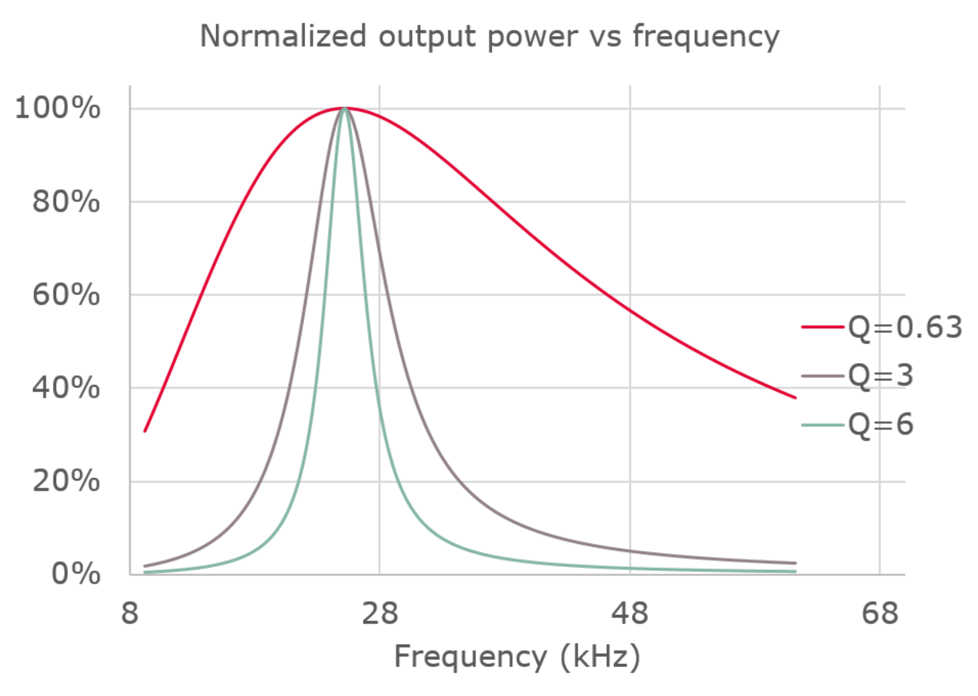


Figure 16 **Output power vs switching frequency for different quality factors.**

In Figure 16, the resistor power dissipation is shown with respect to the switching frequency and at different quality factors, assuming an input voltage amplitude equal to one. As seen in the graph, the output power characteristcs flatten at higher frequency, especially for those with lower Q-factors. This means that for higher values of resistance R, the capability to control of the output power is reduced, and therefore a higher switching frequency is needed to achieve the same output power level. This is one of the reason why the pulse-width modulation stategy is used for very low power settings [12], as shown in the next section.

3.3.2 Operation of half-bridge series resonant converter in pulse-width-modulation scheme

As mentioned before, there can be cases in a multi-hob cooker, where different inverters need different power ratings requiring different operating frequencies. An easier approach to increase the flexibility of control, and still to allow the iso-frequency operation, is to use a pulse-width modulation (PWM) strategy to control the switch. This control method shows that for a given equivalent resonant tank, the maximum output power is achieved when both the half-bridge devices conduct for the same length of time (that is a control signal of 50% duty cycle), while the output power reduces in case of asymmetric conduction times as shown in Figure 17 (duty cycles lower or higher than 50%).

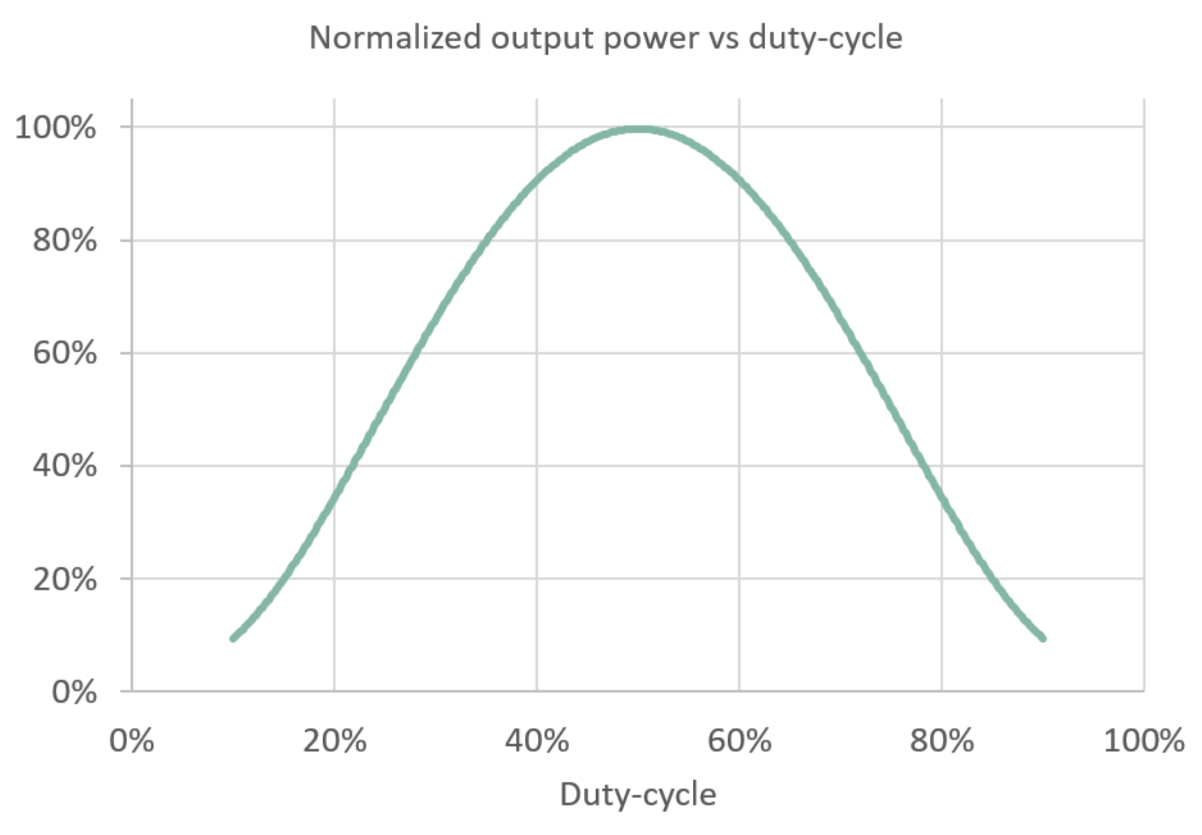


Figure 17 **Resistor real power of a resonant tank vs the duty cycle of the input PWM signal.**

An example of the IGBT switching waveform for a duty cycle (DC) of 20% and 50% for the same resonant tank and switching frequency is shown in Figure 18. As depicted in the graphs by changing the duty cycle, it is possible to achieve considerable load variations with no change of the switching frequency. Figure 18 shows a comparison of the typical waveforms of a half-bridge resonant converter operating with a DC of 20% and 50%, respectively.

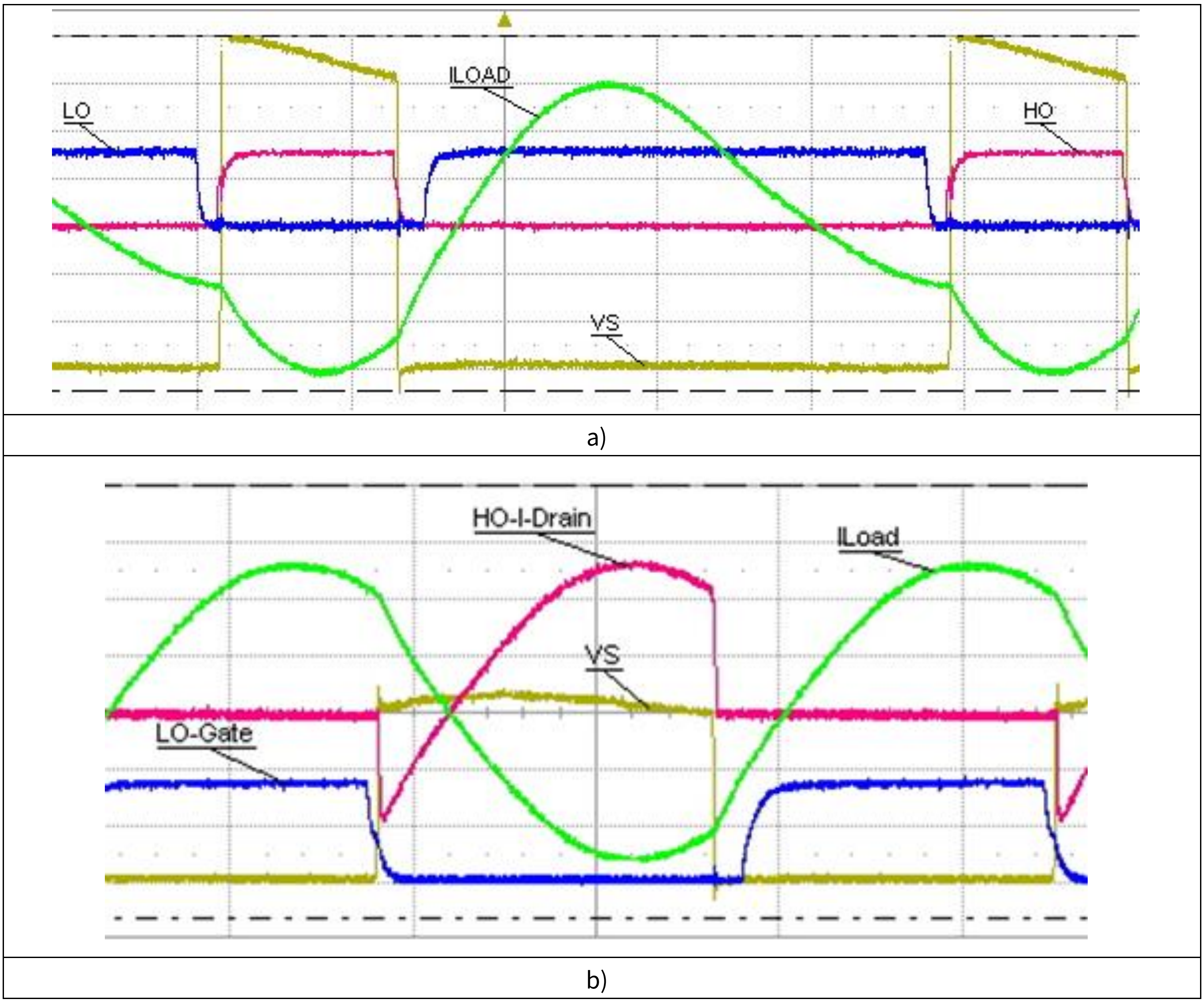


Figure 18 **Comparison of switching waveforms with: a) DC=30%, b) DC= 50%. The considered DC is for the low-side IGBT.**

The PWM control scheme has a clear advantage in terms of ease of control, however, it also has a significant impact on the overall efficiency of the converter, as one of the two power switches no longer operates in soft switching. This leads to an increase of power dissipation, which can be detrimental to converter performance especially at high switching frequencies. For this reason, the PWM strategy is usually used only in extreme cases, when two inverters are required to deliver very different power demands to the respective loads. In such a case, it is convenient to operate the inverters at the lowest output power setting with a low duty cycle, while maintaining the 50% duty cycle control strategy for the inverter with higher power demands.

3.3.3 Advantages and disadvantages of the half-bridge series-resonant converter for induction cooking applications

Some of the advantages of the HBSR topology include:

- Very good power-conversion efficiency due to soft-switching operation of the power switch, at least when PWM control is not used

- Easy to control by changing the switching frequency or by PWM approach
- High efficiency at all output power levels
- 600 V/650 V IGBTs can be used which have a much better conduction vs. switching losses trade-off with respect to IGBTs in higher voltage classes
- Maximum output power not limited by the voltage of the switches

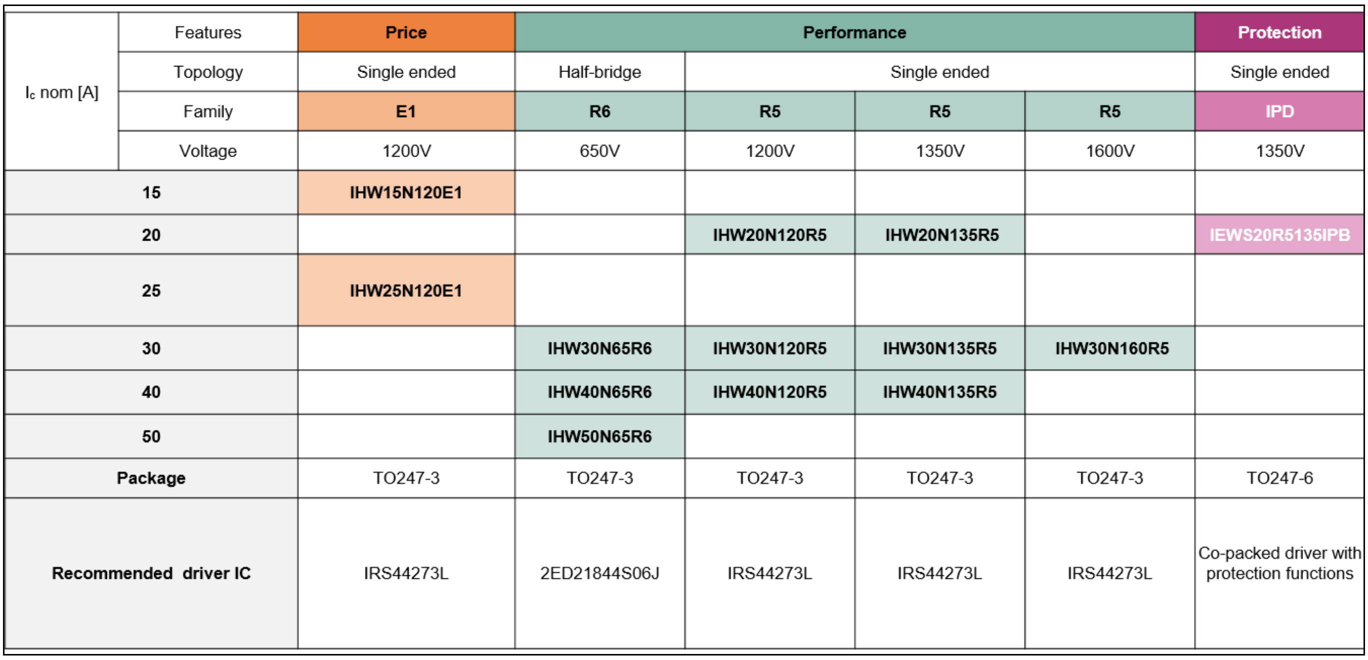
The main drawbacks of the half-bridge topology, with respect to the quasi-resonant topology are related to the higher number of switches:

- More complicated design
- More expensive
- More space is required

The HBSR topology is mostly used in the induction cooking appliances with multiple cooking zones or when the required output power per inverter is above 2.5 kW. Due to the higher cost compared to the quasi-resonant topology, the usage of the half-bridge is mainly restricted to high-end models.

4 Infineon IGBT technologies and products for induction cooking

Infineon provides IGBTs optimized for induction heating applications, both for SEPR and HBSR solutions, for switching frequencies in a range of 20 to 75 kHz. IGBTs for HBSR are suitable for hard- and soft-switching operation in a typical range of switching frequency of 20 – 75 kHz, whilst the high voltage IGBTs are only suitable for resonant switching application, with a range of frequency from 20-50 kHz. IGBTs for induction heating are offered in three series, depending on the main driving factor of the design: price, performance or reliability- driven optimization. A detailed list of the IGBT portfolio for induction cooking is shown in Figure 19[1].



I_c nom [A]	Features	Price	Performance				Protection
	Topology	Single ended	Half-bridge	Single ended			Single ended
	Family	E1	R6	R5	R5	R5	IPD
	Voltage	1200V	650V	1200V	1350V	1600V	1350V
15		IHW15N120E1					
20				IHW20N120R5	IHW20N135R5		IEWS20R5135IPB
25		IHW25N120E1					
30			IHW30N65R6	IHW30N120R5	IHW30N135R5	IHW30N160R5	
40			IHW40N65R6	IHW40N120R5	IHW40N135R5		
50			IHW50N65R6				
Package		TO247-3	TO247-3	TO247-3	TO247-3	TO247-3	TO247-6
Recommended driver IC		IRS44273L	2ED21844S06J	IRS44273L	IRS44273L	IRS44273L	Co-packed driver with protection functions

Figure 19 Infineon portfolio of IGBTs for induction cooking and microwave oven application

The core technology at the basis of Infineon IGBTs for induction cooking is the TRENSHTOP™ reverse-conducting (RC) IGBT technology with monolithically integrated diode. Such a technology is optimized to achieve very low IGBT conduction losses, which represent the largest contribution to the overall power losses of the IGBT. The monolithically integrated diode is designed in such a way that its contribution to conduction losses is minimized, without impacting the performance of the IGBT. Infineon RC IGBTs enable developers of induction heating systems in resonant topologies to design systems with the lowest losses in the application. In this way, the heat transferred to a cooling system can be reduced, resulting in lower cost and effort to design the thermal management of the system.

4.1 IGBT and diode requirements for QR converters

When it comes to the selection of power devices for the QR converter, a designer should pay attention to two main factors: the power dissipation of the device and the maximum electrical ratings. The power dissipation of the device has a direct impact on the design of the cooling system. In the QR topology, as shown in the previous chapter, the losses of the device come mainly from the conduction of the IGBT and the diode, and from the turn-off switching of the IGBT, as shown in Figure 20a. For this reason, an optimal device for a QR topolgy should have very low conduction losses and low turn-off switching losess. This translates directly into requirements for specific device parameters, such as very low V_{CE} saturation voltage and low turn-off energy for the IGBT, and low conduction losses for the diode. The turn-on losses of the diode are usually negligible due to

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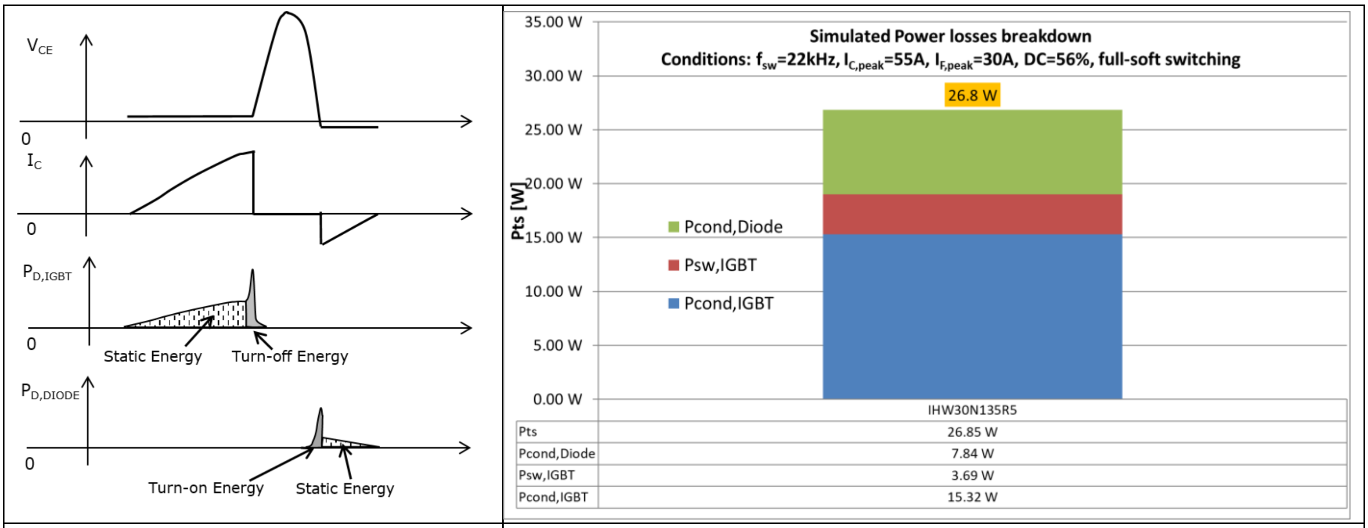
[1] To March 2021.

the low rate of rise of the current at the diode turn-on. In addition, in the QR topology the diode requirements are much less stringent in terms of reduced conduction losses than for the IGBT, since the diode conduction time is much shorter during most of the operation[1]. An example of the power loss split for the Infineon IHW30N135R5 IGBT is given in Figure 20b for typical operating conditions of an inverter that delivers an output power of 2 kW, assuming that the device operates in full soft-switching.



	IHW30N135R5
Pts	26.85 W
Pcond,Diode	7.84 W
Psw,IGBT	3.69 W
Pcond,IGBT	15.32 W

a) b)

Figure 20 **Typical power dissipation contributions of IGBT and diode in a switching cycle of a quasi-resonant converter**

For the maximum electrical ratings, the critical parameters to consider are:

- turn-off current of the IGBT, usually referred as "Turn-off safe operating area." This parameter represents the maximum current that can be switched off safely by the device when it operates in hard-switching conditions, with fast dV/dt transients.
- pulse current of the IGBT (I_{Cpulse}) that is defined as the maximum current-carrying capability of the IGBT, limited only by the junction temperature rise. For Infineon IGBTs for induction cooking, this value is usually the same as the turn-off current.
- pulse current of the diode (I_{Fpulse}) that is defined as the maximum current-carrying capability of the diode, limited only by the junction temperature rise. For Infineon devices for induction cooking, this value is usually the same as for the IGBT.
- device operating junction temperature (T_{vj}) which is usually 175°C for the induction cooking devices from Infineon.
- Device breakdown voltage ($V_{(BR)CES}$), which represents the maximum voltage that is safely sustainable by the device.

In general, a designer should determine the worst-case application conditions, and select the right device according to the parameters above by maintaing a sufficient margin to the maximum limits. Infineon recommends maintaining a margin of at least 20% to guarantee the safest operation of the device over the entire lifetime of the system. Among the parameters mentioned above, $V_{(BR)CES}$ is the most limiting for a QR converter. In fact, as already presented in the previous chapter, the peak voltage across the IGBT in its off-state depends on the specific characteristics of the resonant load and on the value of the bus voltage; it is at any rate

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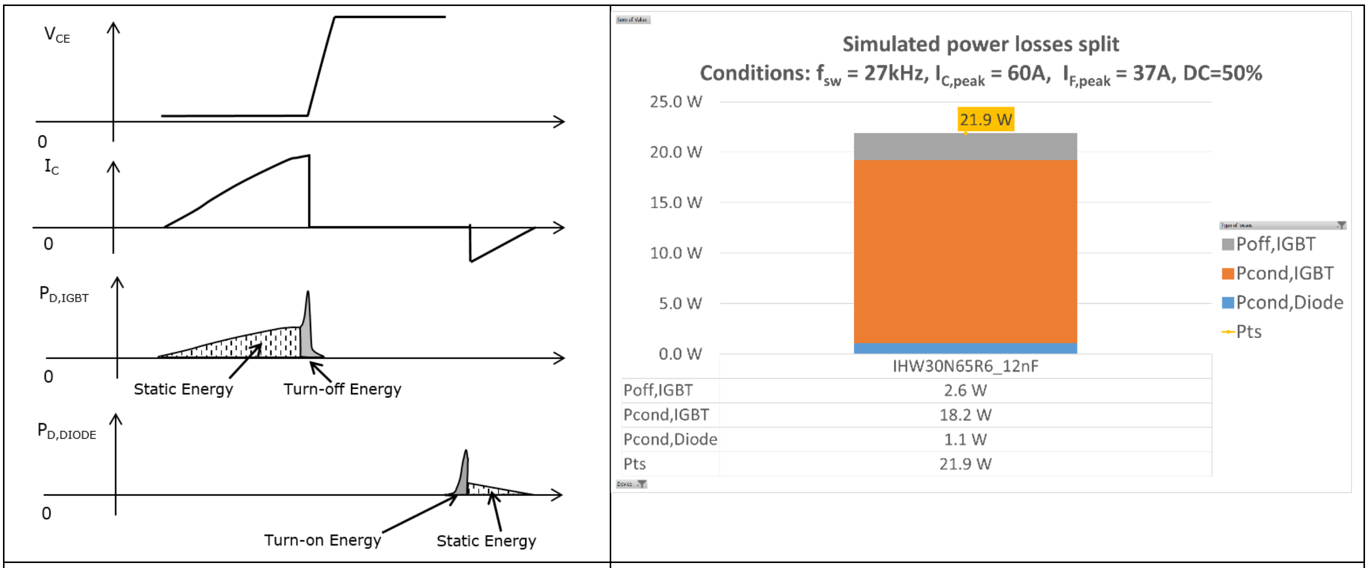
[1] This statement is strongly dependent on the quality of the vessel. The higher is the damping factor of the resonant circuit, which means the higher the resistivity of the load, the shorter is the conduction time of the diode.

proportional to the output power of the inverter. The specific load is not completely predictable during the design phase of the system, as it depends on the specific vessel used, thus it is very important to maintain a sufficient margin of the peak V_{CE} with respect to the breakdown voltage of the device. In this sense, an IGBT with higher $V_{CE,br}$ parameter offers higher resilience against unpredicted load variations and/or input bus voltage fluctuations. That is also the reason for the trend towards higher maximum, sustainable voltages for the IGBTs designed for QR induction cooking applications. Currently, a voltage of 1350 V is usually considered to be sufficient to cover designs with up to 2.2 kW output power per inverter. For higher power, IGBTs with higher voltage classes, like the Infineon IHW30N160R5 are usually necessary.

Beside the IGBTs, in the QR converter, another possible alternative for the power switch is represented by the silicon-carbide MOSFETs, which are available in voltage classes of 1200 V and above. As of today, however, the IGBT still remains the most cost-competitive solution for induction cooking designs based on the single-ended topology.

4.2 IGBT and diode requirements for HBSR converter

For the selection of the right IGBT for a half-bridge converter for induction cooking, similar considerations as for the quasi-resonant converter can be made. Since the device operates in soft-switching turn-on in the half-bridge converter, the main contribution to the losses come mainly from the conduction of the IGBT and the diode, and from the turn-off switching of the IGBT, as shown inFigure 21a. As a consequence, the optimization of the IGBT and diode for this topology follows the same guideline as for the QR topology: very low V_{CE} saturation voltage and low turn-off energy for the IGBT, and low conduction losses for the diode. Also in this half-bridge converter, the turn-on losses of the diode are usually negligible due to the very low rate of rise of the current at the diode turn-on. The main difference between the HB converter and the QR converter is in the optimization of diode turn-off losses. Often in practical cases, the converter can be controlled by means of a pulse-width modulation strategy, for which the zero-voltage swithching turn-on is no longer valid. In case of loss of ZVS turn-on, the diode turn-off losses become relevant for the overall dissipation of the device. Since a compromise between conduction losses and turn-off losses must be found for the diode, the first parameter is generally chosen as design optimization, as the zero voltage switching operation is more frequent, therefore the switching losses are ususally less relevant.



	IHW30N65R6_12nF
Poff,IGBT	2.6 W
Pcond,IGBT	18.2 W
Pcond,Diode	1.1 W
Pts	21.9 W

a) b)

Figure 21 Typical power dissipation contributions of IGBT and diode in a switching cycle of a half-bridge converter

A typical split of power losses for the Infineon IHW30N65R6 is shown in Figure 21b for typical operating conditions of an inverter delivering an output power of 2 kW, assuming that the device operates in full soft-switching for the entire time, and when a total snubber capacitance of 12 nF is used. For maximum electrical ratings, the relevant parameters to consider for selecting the right device are the same as for the quasi-resonant converter. However, in this case, the most critical parameter for the design is represented by the maximum pulse-current capability of the device. The HBSR converter is a current resonant topology, and therefore the peak V_{CE} of the device is constant with the output power. For this reason, the half bridge topology is used mainly in designs with inverter power rating above 2.5 kW.

Worth mentioning for the HBSR is that the IGBT is not the only technology that can be used. Efficient solutions that are based on silicon and silicon-carbide MOSFETs are also available on the market. However, due to the soft-switching operation and the high-current demand, the IGBT remains the most cost-competitive solution for designs up to 100 kHz.

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2.0	January 2014	First release
3.0	20.04.2021	Major updates: - Induction heating section enlarged - Added sections on quasi-resonant converter - Added section on half-bridge converter - Added sections of IGBT and diode requirements for quasi-resonant and half-bridge converters Minor updates: - Updated portfolio in Figure 19
3.01	24.08.2021	Correction of typos and other minor updates

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