Switched-mode converters (one quadrant)

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Abstract
Switched-mode converters are DC/DC converters that supply DC loads with a regulated output voltage, and protection against overcurrents and short circuits. These converters are generally fed from an AC network via a transformer and a conventional diode rectifier. Switched-mode converters (one quadrant) are non-reversible converters that allow the feeding of a DC load with unipolar voltage and current. The switched-mode converters presented in this contribution are classified into two families. The first is dedicated to the basic topologies of DC/DC converters, generally used for low- to mid-power applications. As such structures enable only hard commutation processes, the main drawback of such topologies is high commutation losses. A typical multichannel evolution is presented that allows an interesting decrease in these losses. Deduced from this direct DC/DC converter, an evolution is also presented that allows the integration of a transformer into the buck and the buck–boost structure. This enables an interesting voltage adaptation, together with a galvanic isolation directly integrated into the converter. The second family is related to DC/DC converters with an intermediary AC stage. Such structures include middle-frequency transformers as described above, and offer reduced commutation losses thanks to natural soft commutation conditions, sometimes reinforced by the insertion of LC components or active devices. This allows high switching frequencies, and then a reduction of the size and weight of such applications.

1 Introduction

Switched-mode converters are DC/DC converters, dedicated to the supply of DC loads with a regulated output voltage and protections against overcurrents and short circuits. These converters are generally fed by an AC single-phase or three-phase network, as presented in Fig. 1.

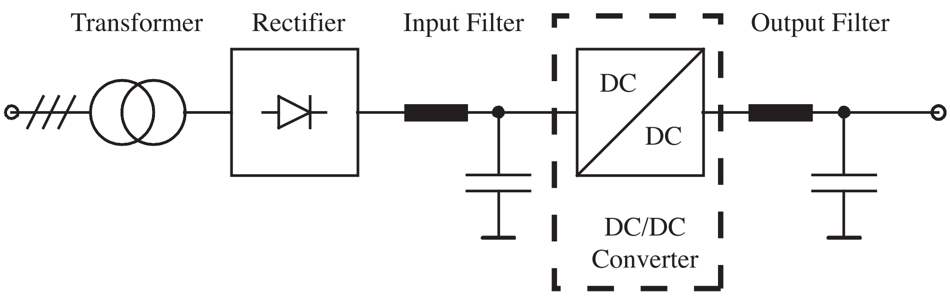


Fig. 1: General scheme of a DC supply

From an AC network, a conventional conversion chain is made of an input transformer dedicated to voltage/current adaptations, offering also galvanic insulation of its secondary side with respect to its primary. Then, a non-controlled rectifier (diode rectifier) realises the conversion from AC to DC. The voltage generated is not regulated, and presents a strong ripple around its averaged value. An LC input filter is introduced in such a way that it cancels or minimizes this voltage ripple. Moreover, this input filter acts as a voltage source for the DC/DC converter needed to regulate the voltage applied to the load. Sometimes, the DC/DC converter has to limit the current absorbed by the load. In a lot of applications, an output filter is required to filter the output voltage ripple.

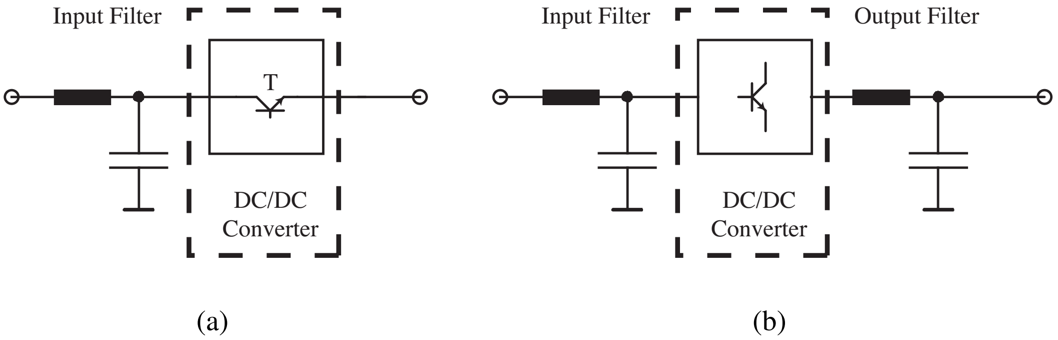


Fig. 2: DC/DC converters: (a) Linear converter. (b) Switched-mode converter.

Generally speaking, one can divide DC/DC converters into two families of converters, as shown in Fig. 2. For low-power applications, the DC/DC converter can be realized with transistors that are series connected to the main current flow Fig. 2(a). Such transistors are driven in their linear mode. They have to support the difference of potential between the non-regulated voltage from the input filter and the voltage requested by the load. Although such a solution offers a high dynamic for the output voltage regulation and a poor output voltage ripple, its efficiency is low since losses are proportional to the current needed by the load and the non-zero voltage the transistor has to maintain. This solution is mainly used in low-power applications.

The only way to obtain a high efficiency in mid- or high-power applications is the use of switched-mode DC/DC converters Fig. 2(b), using transistors as interrupts, switching between their off-state and their saturated mode. Here, losses are related only to the switching losses and low conduction losses, ensuring a higher efficiency and thus a higher power density. This is the main advantage of switched-mode DC/DC converters, even if the commutations lead to non-negligible EMC emissions and if an output filter is needed to lower output voltage ripple.

This contribution concerns switched-mode DC/DC converters. Depending on the application, these converters can have one, two, or four quadrants depending on their topologies, and allow the reversal of the sign of the output voltage and/or the output current. We shall focus our presentation on one-quadrant switched-mode converters that allow energy flows in only one direction, i.e., from the main energy source to the load, this means the supply of a load with only positive voltage and current. Such converters can be classified into two families:

- Direct DC/DC converters: one single stage is used to adjust voltage levels from the input to the output. These converters are generally obtained with one switching cell. Three of them will be presented:
  - buck converter (or step-down converter),
  - boost converter (or step-up converter),
  - buck–boost converter (or step-up/down converter).

  From these basic topologies, we shall go on to explain how to increase the efficiency of such solutions thanks to the multichannels technology, allowing the converters to work under soft–switching conditions. Some converters can be adjusted to allow sinusoidal current absorption on the feeding AC network (Fig. 1). This will be illustrated with the boost converter. Moreover, it is possible to add transformers in the switching cell of the buck converter and the buck–boost converter, even if such topologies have not been specifically designed to generate AC voltages or current. Two other topologies offer DC/DC conversion together with galvanic insulation and additional voltage/current adaptation thanks to their internal transformers:

  – forward DC/DC converter,
  – flyback DC/DC converter.

– DC/DC converters with an intermediary AC stage: this conversion chain is described in Fig. 3. The first stages are similar to those described in Fig. 1. From the input filter, a voltage source for the following converter, a DC/AC converter feeds an alternative intermediary stage. Then depending on the application, an LC resonant circuit is added to enable soft-switching conditions for the DC/AC converter, in order to maximize its efficiency. A middle-frequency transformer can then be inserted to allow voltage adaptation and galvanic insulation. In some applications, this should avoid the use of the input transformer. Then, a rectifier (generally non-controlled in one-quadrant applications) feeds the output DC stage with an output filter. The regulation of the output voltage is achieved by the control of the DC/AC converter; particular attention has to be given to the middle-frequency transformer where DC components must be avoided. We shall describe

  – the possible topologies for the DC/AC converter,
  – the possible topologies for the rectifier,
  – controls for the DC/AC converter that allow the regulation of the output voltage,
  – the use of resonant LC circuits in order to allow soft-switching conditions for the DC/AC converter.

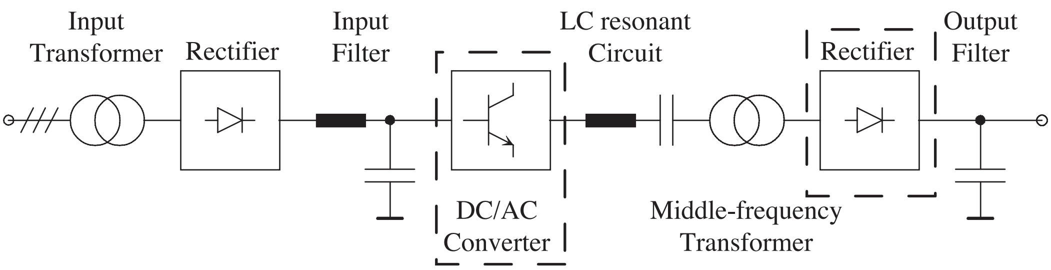


Fig. 3: DC/DC converter with AC intermediary stage

2 Direct DC/DC converter

2.1 Main structures

In this section, we describe the three main topologies representing the basis of DC/DC conversion [1–3]. They are presented in Fig. 4.

The first structure [Fig. 4(a)] is the buck converter, defined as an interface between a DC voltage source (the input filter in Fig. 1) and a DC current source, the inductor L_s of its output filter. This structure is obtained with only two interrupts: a transistor T and a diode D. Each controlled transition of T induces the spontaneous commutations of the diode. As a result, the output voltage U_s is made of two distinct levels: U_e and 0 V. As the LC filter is designed to lower voltage and current ripple by the commutations of the switches, the voltage U_{so} applied to the load is the averaged value of U_s. Such a converter is then a step-down converter, where U_{so} is always lower than the input voltage U_e. Each of the two interrupts must be able to maintain the voltage U_e in their off-state. They also need to support at least the current needed by the load in their on-state.

The structure in Fig. 4(b) is the boost converter, defined as an interface between a DC current source made of an inductor L_e (the capacitor of the input filter in Fig. 1 can then be avoided), and a DC voltage source, the capacitor C_s of its output filter. As for the buck, a transistor T and a diode D

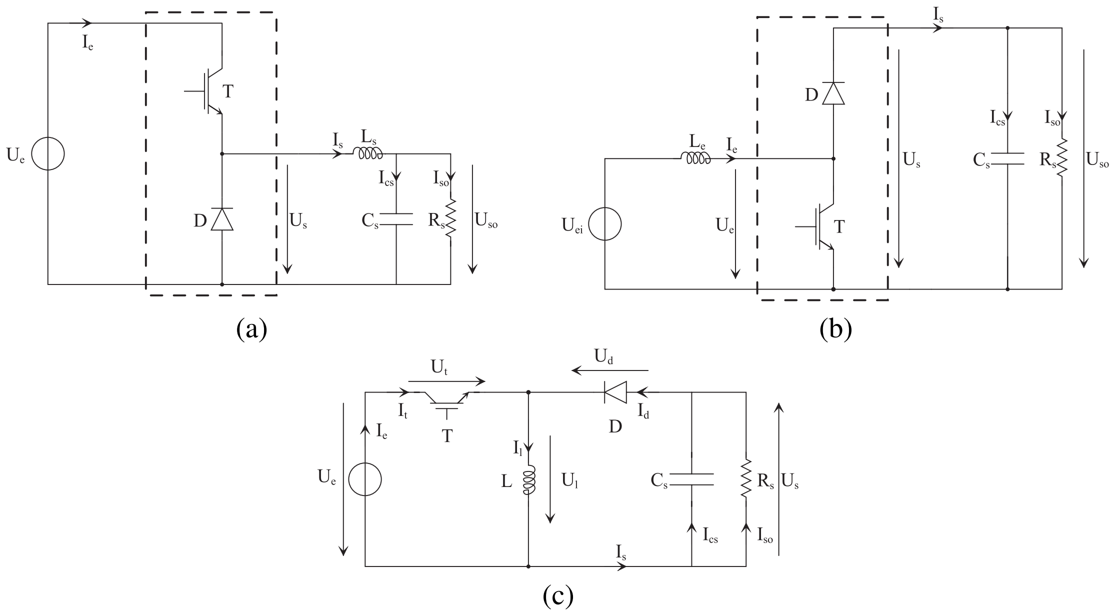


Fig. 4: Main structures: (a) Buck converter. (b) Boost converter. (c) Buck–boost converter.

form the commutation cell. Each spontaneous commutation of D is linked to the controlled transitions of T. When the transistor T is in its on-state, the intermediary voltage U_e is 0 V. As the voltage across the inductor L_e is U_ei and is positive, the current I_e increases. When the transistor T is turned off, the diode spontaneously switches on. The voltage across the inductor L_e is the difference between the input voltage U_ei and the output voltage U_so. This phase must correspond to a decrease of the current I_e to keep control of the structure. The voltage across L_e must then be negative, this means that the output voltage U_so must be higher than the input voltage U_ei. Such a converter is then a step-up converter. Each of the two interrupts must be able to maintain the maximum output voltage U_so while in their off-state. They also need to support at least the maximum current in L_e.

The last structure Fig. 4(c) is the buck–boost converter, defined as an interface between a DC voltage source (the input filter in Fig. 1), and another DC voltage source, the capacitor C_s of its output filter. As a direct converter can interface only electrical sources of opposite nature, an intermediary inductor L is added. This converter is made of one switching cell, consisting of a transistor T and a diode D. Each spontaneous commutation of D is triggered by the controlled transitions of T. It should also be noted that the two interrupts can be simultaneously in their on-state. Then, when T is in its on-state, the diode acts as an open switch. The input of the converter and its output are isolated, the voltage across L is positive, and the current in it increases. When the transistor T is off, the diode is in its on-state. The input of the converter and its output are still isolated. Part of the energy stored in the inductor L is transmitted to the output stage. Regarding the conventions chosen for the sign of the output voltage, the voltage across L is then negative, and the current I_l is decreasing. The energy transfer between the two voltage sources is then not continuous. Energy is transferred from the input to the inductance in a first step of the switching period of the converter. In a second step, this energy flows from the inductor to the output stage. The particularity of the buck–boost converter is that each of the two switches has to be able to maintain the sum of the voltages U_e and U_s in their off-state.

Whatever the topology, two main parameters define their behaviour as shown in Fig. 5.

The first parameter is the duty cycle D, defined as the ratio between the conduction time of the transistor T and the switching period of the topology. This duty cycle can vary from 0 (transistor always in its off-state) to 1 (transistor always saturated).

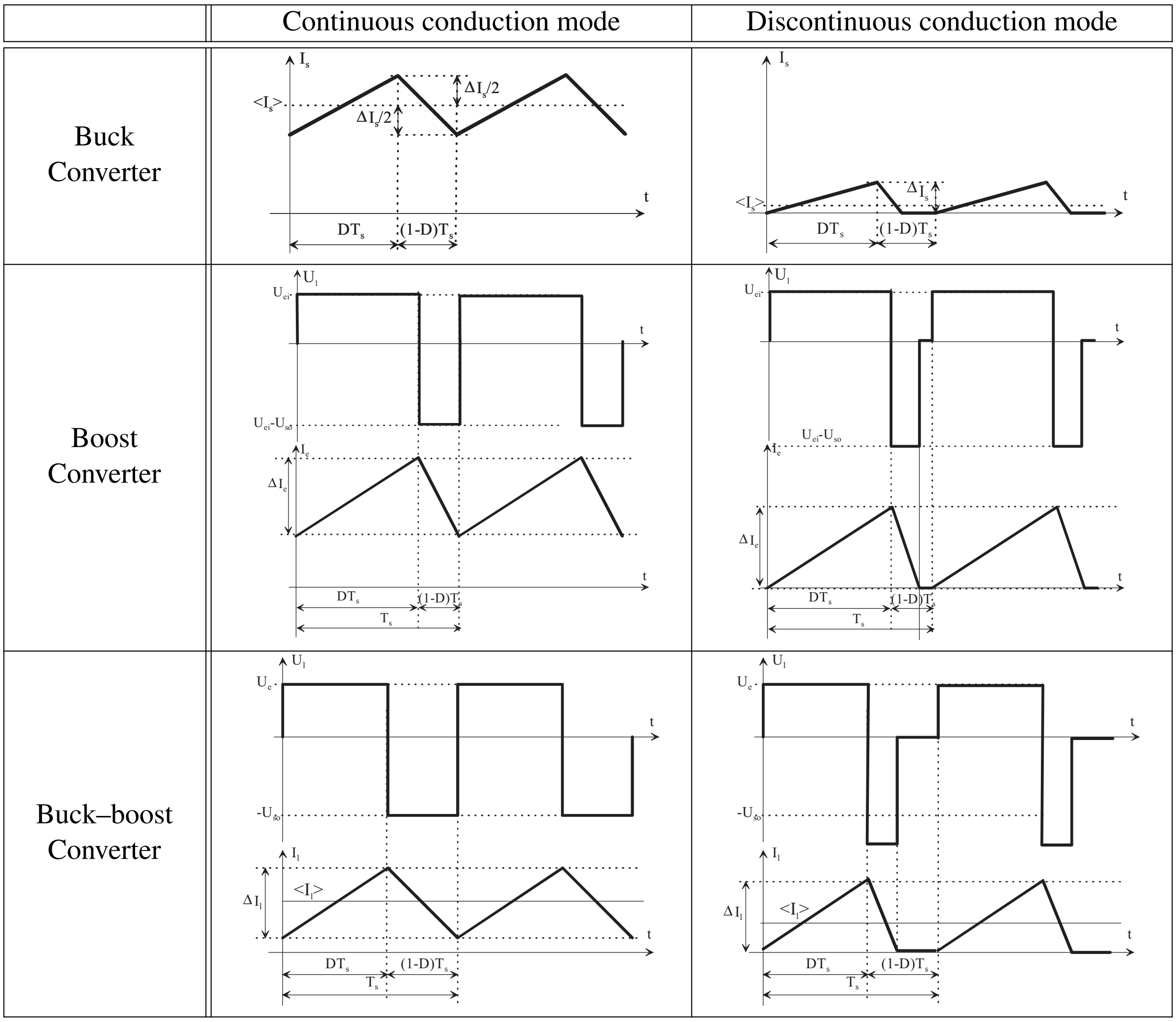


Fig. 5: Main characteristics of direct DC/DC converters

The second parameter is the currents circulating through the inductor (L_s for the buck, L_e for the boost, L for the buck–boost). Two main cases can be considered:

- Continuous conduction mode: where the current in the inductor is always different from zero. There, each commutation of the diode is trigged by the controlled commutations of the transistors, in a hard-commutation switching process.
- Discontinuous conduction mode: the current in the inductor has a strong ripple, non-negligible in terms of its averaged value. As a consequence, each off-switching of the diode is due to the natural decrease of the inductor current to 0, while the transistor is not immediately switched on. Moreover, the transistor is switched on while the currents in the structure are zero. This represents a Zero-Current Switching (ZCS) of the transistor. Coupled with the natural blocking of the diode, this enables reduced switching losses compared to the continuous conduction, with increased efficiency.

The relations between the output voltage and the input voltage are mainly defined by the duty cycle, but are strongly influenced by the conduction mode of the converter, as illustrated in Fig. 6.

In the continuous conduction mode, the relations between output and input voltages are defined only by the duty cycle. One can then recognize the step-down nature of the buck converter, the step-up

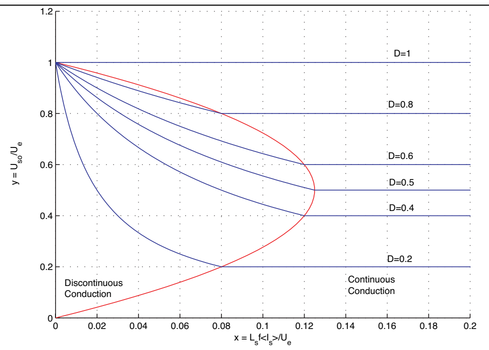
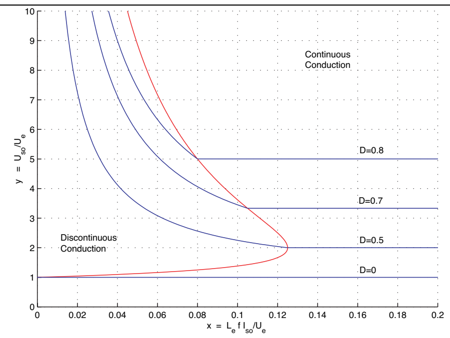
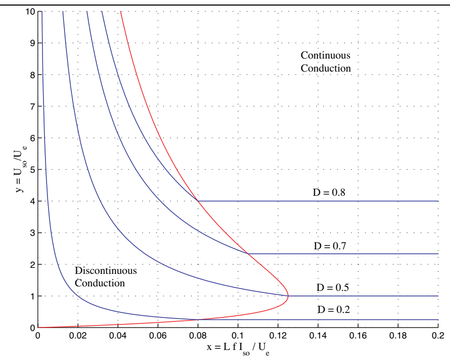
<table>
<tr><th></th><th>Main equations</th><th>Output characteristics</th></tr>
<tr><td>Buck converter</td><td>Continuous conduction mode
$$\frac{<U_\mathrm{s}>}{U_\mathrm{e}} = D$$
Discontinuous conduction mode
$$\frac{<U_\mathrm{s}>}{U_\mathrm{e}} = \frac{1}{1+\dfrac{2L_\mathrm{s}f<I_\mathrm{s}>}{D^2U_\mathrm{e}}}$$</td><td></td></tr>
<tr><td>Boost converter</td><td>Continuous conduction mode
$$\frac{<U_\mathrm{s}>}{U_\mathrm{e}} = \frac{1}{1-D}$$
Discontinuous conduction mode
$$\frac{<U_\mathrm{s}>}{U_\mathrm{e}} = \left(1+\frac{U_\mathrm{ei}}{2L_\mathrm{e}f<I_\mathrm{s}>}D^2\right)$$</td><td></td></tr>
<tr><td>Buck–boost converter</td><td>Continuous conduction mode
$$\frac{<U_\mathrm{s}>}{U_\mathrm{e}} = \frac{1}{1-D}$$
Discontinuous conduction mode
$$\frac{<U_\mathrm{s}>}{U_\mathrm{e}} = \left(\frac{U_\mathrm{e}}{2Lf<I_\mathrm{s}>}D^2\right)$$</td><td></td></tr>
</table>

Fig. 6: Output characteristics of DC/DC direct converters

nature of the boost converter, and the step-up/down nature of the buck–boost, when D varies from 0 to 1. If the transistor and the diode of each structure suppose no losses in their on-state, and if the switching processes are considered instantaneous, then the output voltage is dependent only on the duty cycle as shown on the output characteristics.

It is different in the discontinuous conduction mode, where the relationship between output and input voltages is defined not only by the duty cycle, but also by the value of the inductor and the switching frequency, and by the output current. In particular, the duty cycle of each converter has to be adapted for each decrease of the output current in order to maintain constant the output voltage. One can note that in the case of low output currents for the boost and the buck–boost converters, the output voltage can reach huge values, even for reduced values of the duty cycle.

2.2 Multichannel converters

Even if the control of the DC/DC converters we have described is more complex in the case of the discontinuous conduction mode, such a working mode is interesting because it offers increased efficiency.

However, a DC/DC converter working in the discontinuous conduction mode offers a poor averaged value of the generated currents for the required current ripple. A way to solve this problem is to use the technology of multichannel converters, as described in Fig. 7 for a four-channel buck converter.

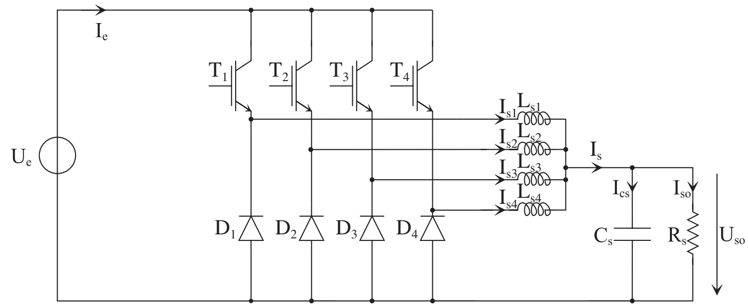
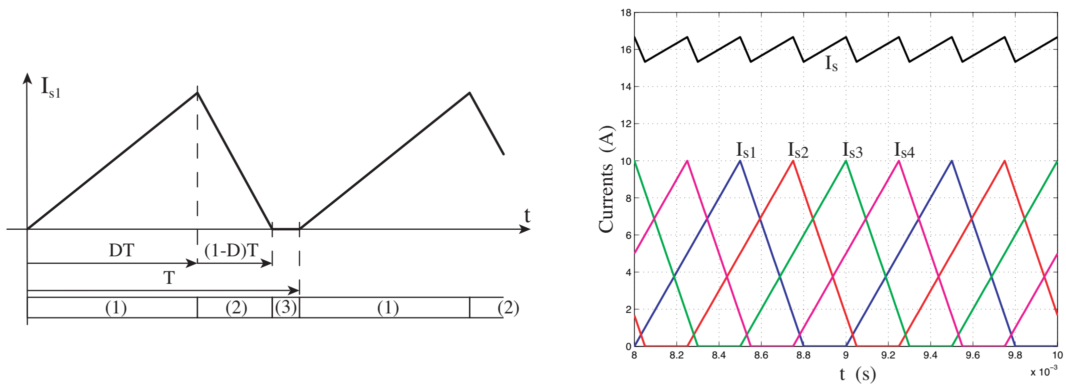


Fig. 7: Interlaced channels buck converter

In this example, four elementary buck converters are associated. Each of them is fed by a common voltage source. Each elementary converter is connected to an inductor that acts as a current source at the switching phenomenon scale. The currents in each inductor are then summed, and injected into the output capacitor and the load. Each leg works in the discontinuous conduction mode, as described in Fig. 8(a) for the elementary converter made of T_1 and D_1.



(a) (b)

Fig. 8: Discontinuous conduction mode: (a) Current in channel 1. (b) Current for a four-channel buck.

In phase 1, the transistor T_1 is on. As the current is zero at the beginning of this phase, switching losses are negligible. The voltage across L_{s1} is positive, and the current I_{s1} increases. At the end of phase 1, the transistor T_1 is switched off. This induces the spontaneous switching-on of the diode. In phase 2, the voltage across L_{s1} is negative, and the current I_{s1} is decreasing. When it reaches zero, the diode spontaneously switches off, while the transistor T_1 is still in its off-state. This switching-off of the diode generates fewer losses than the case where the diode switches off because of the switching-on of the transistor (continuous conduction mode). During phase 3 the transistor and the diode are in their off-state, the current in L_{s1} is zero, until the end of the switching period.

Each of the three other elementary converters acts the same way. However, for a four-channel application, each leg is phase-shifted on a multiple of the fourth of the switching period. The main result is presented in Fig. 8(b). The currents in the inductors have the same averaged value and the same ripple, and are phase-shifted. As a result, the total current I_s is the sum of all the averaged values of the currents generated by each leg, with a reduced ripple due to the phase-shift of the elementary converters.

The main interest of this technology is to propose a high-efficiency power converter (due to reduced switching losses obtained with the discontinuous conduction mode), with reduced inductor (to obtain discontinuous conduction), and a limited output current ripple.

2.3 Power factor correction

As described in the general introduction (Fig. 1), DC/DC switched-mode converters are generally fed by an AC network via a non-controlled rectifier. This is the case of the following structure in Fig. 9, where a boost converter is fed by a diode rectifier.

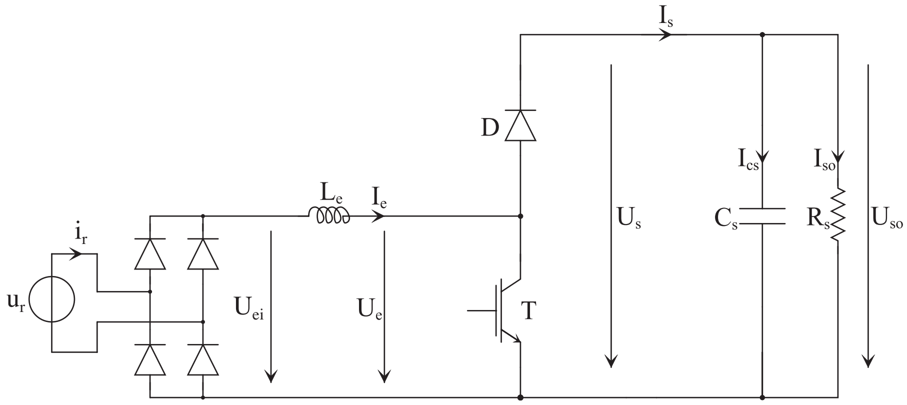


Fig. 9: Boost converter for UPFC

Following the general design rules for power electronics, if the AC network is considered as an alternative voltage source, then the secondary of the diode rectifier must be homogenous to a current source, that is the input inductor L_e of the boost converter, with no need for an additional capacitor. The voltage generated by the diode rectifier is not regulated; the boost converter allows the regulation of the DC voltage U_s. As mentioned in Section 2.1, the strict requirement to keep the control of the boost converter is related to voltage levels: the output voltage must always be higher than the input voltage U_{ei}. The input voltage u_{ei} is not constant, but is the rectified alternative voltage u_r, as shown in Fig. 10.

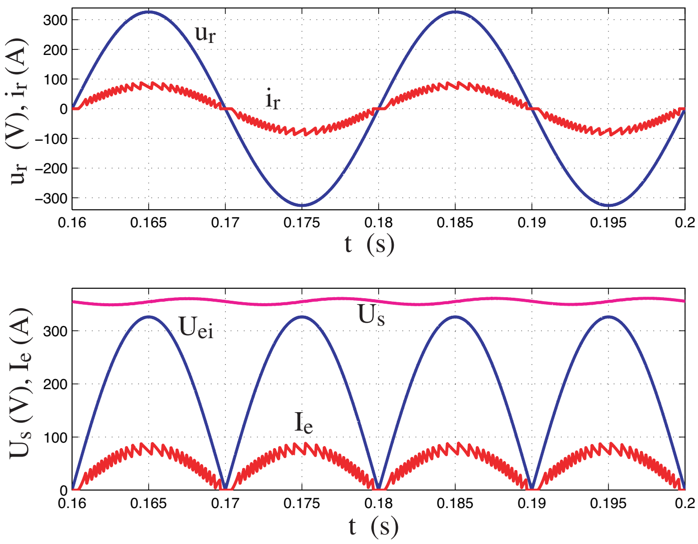


Fig. 10: Typical waveforms

In cases where U_{ei} reaches values that are higher than U_{s}, then the control of the current in the inductor is lost, because there is no way to obtain the conditions to block the diode, which will stay in its on-state until U_{ei} becomes lower than U_{s}. There, the boost topology is transparent, and acts as a 'simple' LC filter, with no regulation of the output voltage U_{s}.

The case presented in Fig. 10 is related to an output voltage U_{s} higher than the rectified voltage U_{ei}. Then control of the current in the inductor L_{e} can always be effected following two criteria:

– regulation of the output voltage, that has to be kept as stable as possible, with the help of the output filter C_{s};
– the harmonic content of the current i_{r} absorbed on the AC network must be as close as possible to the 'pure' sine-wave, with a unit power factor.

To match this last criterion, an alternative component deduced from u_{ei} is superimposed onto the current I_{e} needed to maintain the voltage U_{s} to the given reference. The boost input current is then made of three main components:

– an averaged value: needed to maintain the voltage U_{s} to the given reference;
– a low-frequency ripple: defined by that of the rectifier voltage U_{ei};
– a high-frequency ripple: defined by the switching frequency of the boost converter.

It is evident that the addition of a low-frequency ripple to the current I_e has an impact on the sizing of the output filtering capacitor C_s, that has not only to filter current components linked to the switching frequency of the converter but also has a low-frequency component.

But the main advantage of this kind of control is that, if the low-frequency ripple of the current I_{e} is made with semi-sinusoids, the AC current i_{r} will be close to the 'pure' sinusoid [4]. Its harmonic content will be a fundamental wave, with the same frequency as the AC network. Then, sub-harmonics will be defined by the switching frequency of the boost converter, much higher than that of the AC network. In some cases, these harmonics can be filtered with high-frequency filters. Attention must be given to each zero crossing of the voltage u_{r}, and as a consequence of the voltage U_{ei}. In these special cases, it is difficult to control the current I_{e} that introduces harmonic distortion for the AC current i_{r}.

This principle can also be realized with a buck–boost converter, with the advantage that the current in its inductor can always be controlled, whatever the values of the input and the output voltage may be. The main drawback is that the input current of this converter is strongly discontinuous, with the consequence of strong current ripple at the buck–boost switching frequency level on the AC current i_{r}.

2.4 DC/DC direct converters with transformers

2.4.1 General comments

As described in the general introduction and illustrated in Fig. 1, a transformer is generally inserted between the AC network and the non-controlled rectifier that feeds the DC/DC converter. This transformer is needed to adapt voltage levels, and to ensure galvanic isolation from the alternative network. If the transformation ratio is one of the main parameters for sizing the input transformer, particular attention must be given to other parameters that define the weight and the size of this transformer, and particularly the working frequency.

For an input transformer, this frequency is defined by the frequency of the AC network, generally 50 Hz or 60 Hz for industrial networks, 400 Hz for aerospace networks, etc. The weight and the size and the input transformer are proportional to the inverse of the frequency. It is interesting to increase the working frequency, to lower the dimensions of the transformer. The only solution for the general scheme (Fig. 1) is to modify this conversion chain by inserting the transformer into the DC/DC converter, where the switching frequency is much higher than the AC network frequency [2], [3], [5].

As the DC/DC converters we describe here are direct converters, with no intermediary AC stage, particular attention must be given to the feeding of the transformer, to ensure that it does not contain a DC component.

2.4.2 Flyback converter

The buck–boost converter Fig. 4(c) introduced in Section 2.1 is a converter that interfaces two DC voltage sources. To succeed, an intermediary inductor L is needed. The energy transfer between the input voltage source and the output one is made sequentially: first, the transistor T is switched on, and energy is stored in the inductor. Then the transistor is switched off, the diode D is switched on, and part (or the totality) of the stored energy is transferred from the inductor to the output stage of the converter.

In steady-state working mode, the averaged current into the inductor is constant. This means that the averaged voltage across the inductor L is zero. The voltage across L is purely alternative. It is equal to the mains feeding positive voltage U_{e} while the transistor T is on, and equal to the negative output voltage $-U_{\mathrm{s}}$ when the diode is on. The reader can refer to Fig. 5 to see an illustration of the voltage across the inductor of the buck–boost converter.

As the voltage across L is purely alternative, an evolution of the buck–boost converter is to replace the inductor L with a transformer, whose magnetizing inductance will act as the initial L. This leads to the flyback converter presented in Fig. 11.

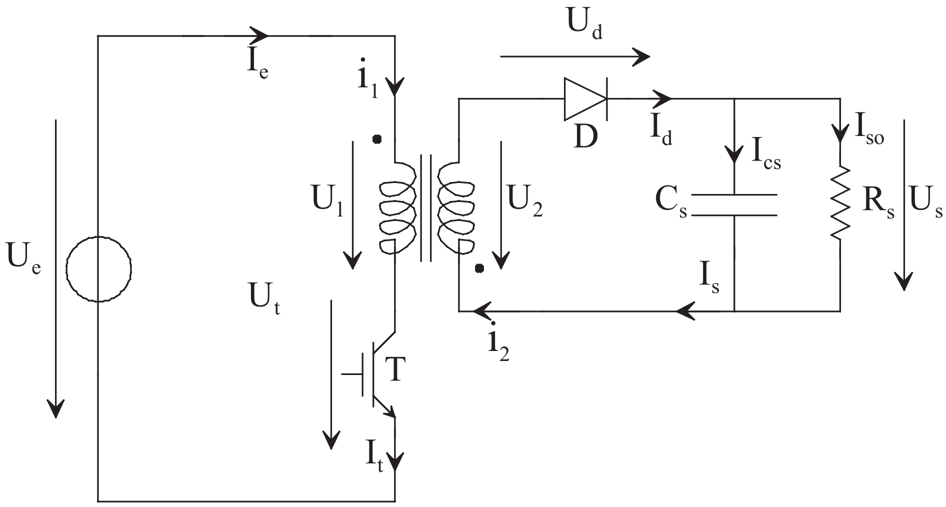


Fig. 11: Flyback converter

Such a converter acts exactly as a buck–boost converter, and two working modes can be exploited: the continuous conduction mode, and the discontinuous conduction mode. The typical waveforms of each of these two working modes are presented in Fig. 12.

The voltage U_1 at the primary side of the transformer is equal to the positive voltage U_{e} when the transistor T is switched on. The voltage U_2 at the secondary side is then mU_{e} (where m is the voltage transformation ratio of the transformer), and is negative: the diode D is then in its off-state. There is then no current on the secondary winding of the transformer, and the transformer can be modelled by its magnetizing inductor series connected to the transistor T. The primary current I_1 then increases, and magnetic energy is stored in the core of the transformer.

When the transistor T is turned off, the continuity of the magnetic flux in the transformer core must be obtained. This is realized thanks to the diode D that turns on, enabling the commutation of the magnetizing current from I_1 to I_2 as there can be no current on the primary side of the transformer. This

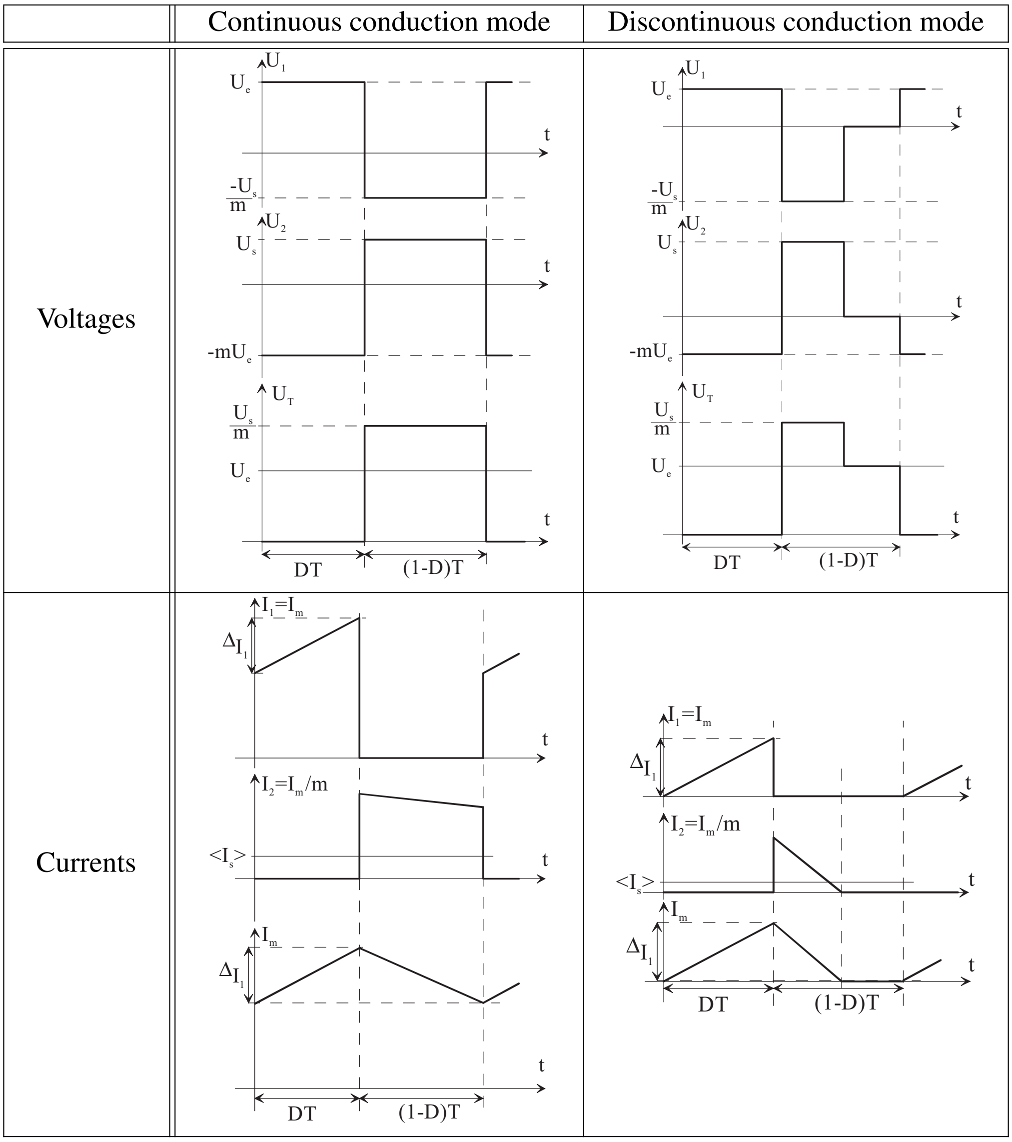


Fig. 12: Flyback converter: typical waveforms

can be modelled by its magnetizing inductor series connected to the diode D. The secondary current then decreases, and the magnetic energy stored in the core of the transformer is transferred to the output stage of the converter. Then, two cases can be considered:

– Continuous conduction mode: the transistor T is turned on while the current I_2 is not yet zero. Then, as a natural switching process strictly equivalent to that of a buck–boost converter, the diode D spontaneously switches off and the current I_2 becomes zero. The magnetizing current is then commuted from the secondary side to the primary side, and starts to increase again.
– Discontinuous conduction mode: the current I_2 reaches zero before turn-on of the transistor T. The transformer is then completely demagnetized. As long as the transistor is not switched on, all currents remain null. This working mode is also strictly equivalent to that of a buck–boost converter in discontinuous conduction mode.

It has to be noted that, whatever the working mode, the main transistor T has to be able to maintain in its on-state a combination of the input and the output voltage ($U_\mathrm{e} + \frac{1}{m}U_\mathrm{s}$).

All the equations defined for the buck–boost converter and the output characteristics (Fig. 6) are still valid for the flyback converter where the transformation ratio m of the transformer must be taken into account. Then, using the duty cycle D, for a switching frequency f and magnetizing inductor L, one can establish that:

- Continuous conduction mode: $U_\mathrm{s} = m\frac{D}{1-D}U_\mathrm{e}$. The ratio between the output and the input voltage is a function of D (that can be adjusted by the control from 0 to 1) and m (constant).
- Discontinuous conduction mode: $U_\mathrm{s} = DU_\mathrm{e}\sqrt{\frac{R}{2Lf}}$. As for the buck–boost, the output voltage is strongly affected by the load that the control must take into account. One can note that the transformation ratio m does not appear in this working mode. The purpose of the transformer is to allow galvanic isolation.

Finally, the following comments can be made:

- The use of a transformer as an intermediary storage device (particularly in the continuous conduction mode) can lead to increased weight and volume.
- The currents at the primary and secondary sides of the transformer have strong variations during commutations. Coupled with leakage inductors of the transformer, this leads to important voltage constraints on the active components.

2.4.3 *Forward converter*

A way to solve the problem of the weight and volume of the transformer in a flyback converter, where it has to be designed to store the energy needed by the load, is to insert the transformer in a structure where no energy storage is needed, and where the transformer could be considered as transparent regarding the power needed by the load. This leads to the forward converter Fig. 13, directly descended from the buck converter.

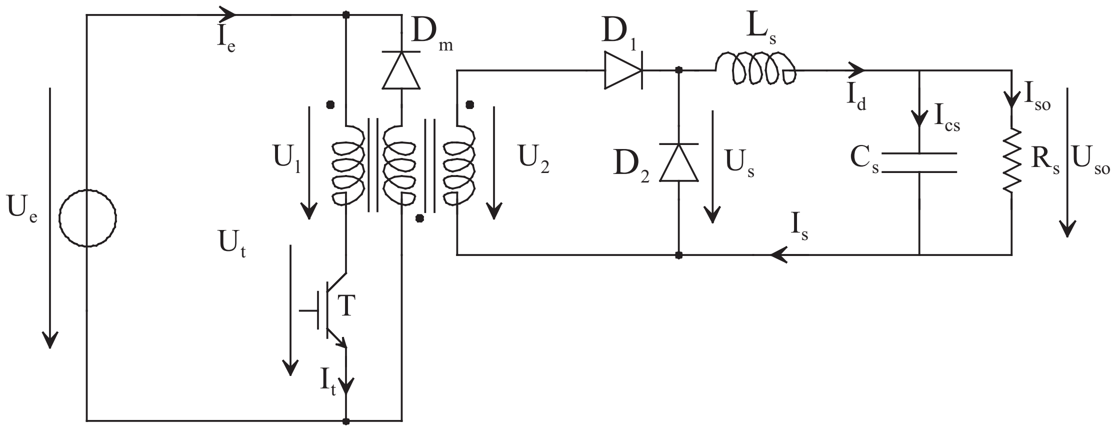


Fig. 13: Forward converter

Considering the buck converter [Fig. 4(a)], the voltage U_s varies between two levels U_e (transistor turned on) or 0 V (diode turned on), around an averaged value $\langle U_\mathrm{s}\rangle$ defined by the duty cycle D: $\langle U_\mathrm{s}\rangle = DU_\mathrm{e}$.

In the forward converter, the primary winding of a transformer is inserted in series with the main transistor. The secondary side of the transformer is connected to the output of the converter, where the diode D_2 plays exactly the same role as the free-wheeling diode D of the buck converter. Another diode D_1 has been inserted to avoid reverse current into the secondary winding of the transformer.

A third winding is added, activated or not, thanks to the diode D_{m}. This winding is necessary because of the voltage applied to the primary side. Initially, the intention is to apply $U_1 = U_e$ when the transistor is on, and $U_1 = U_2 = 0$ V when the diode D_2 is on during the free-wheeling phase of the converter. In this case, the averaged value on the primary voltage is non-zero. The magnetizing current of the transformer could then reach high values incompatible with the sizing of the various components of the structure, together with the saturation of the magnetic core of the transformer. The third winding is then designed to assure the full demagnetization of the transformer to be completed at each end of the switching period of the converter.

Typical waveforms for such a converter are given in Fig. 14.

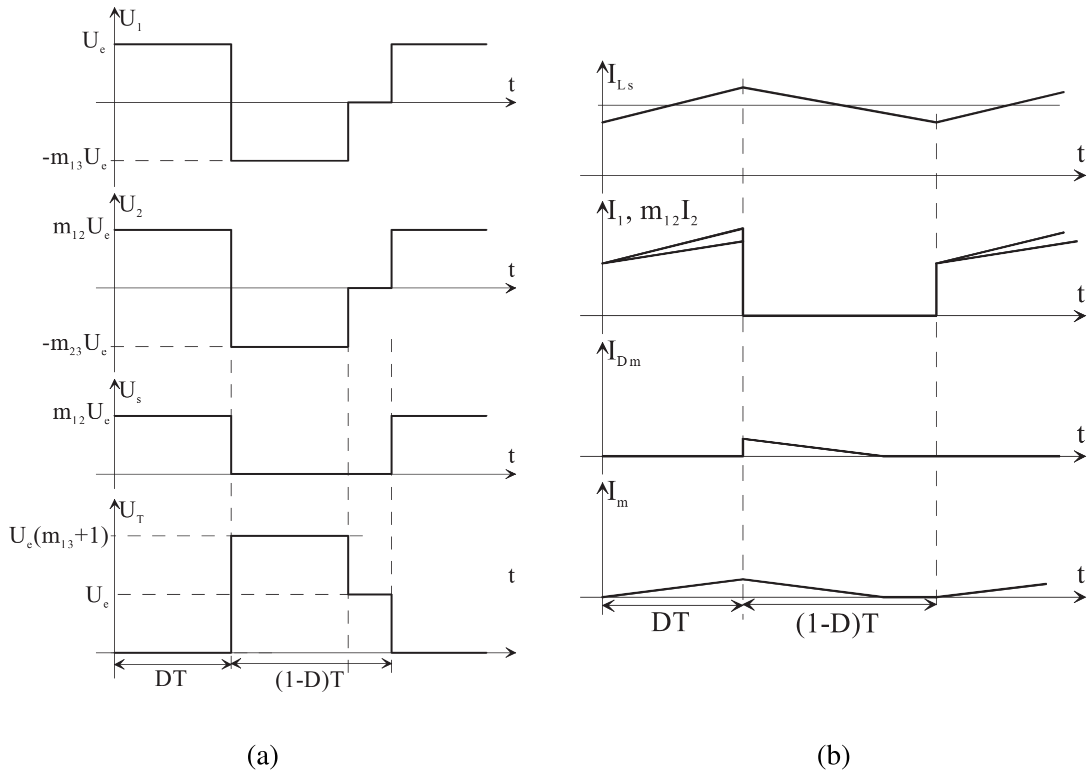


Fig. 14: Forward converter: typical waveforms. (a) Typical voltages. (b) Typical currents.

The voltage U_1 at the primary side of the transformer varies exactly as the voltage U_{s} of the buck converter. It is U_{e} when the transistor T is turned on. As a result, the voltage U_2 at the secondary side varies as U_1, affected by the transformation ratio m of the transformer: $U_2 = mU_1$. From the secondary side of the transformer, the voltage U_2 has to be higher than the output voltage U_{so} to be controlled, as for a buck converter. Then, a current circulates through the secondary winding of the transformer to the output stage. The current at the primary side of the transformer is the sum of two components: the current mI_2 due to the transform ratio, plus the magnetizing current.

When the transistor is turned off, no current can circulate into the primary side of the transformer. To allow the continuity of the magnetic flux in the core of the transformer, the diode D_{m} spontaneously switches on. The voltage $-U_{\mathrm{e}}$ is then applied to the demagnetization winding. This causes the magnetizing current to decrease. During this phase, the voltages U_1 and U_2 are negative. The diode D_1 does not allow any reverse current into the transformer, while the transistor has to support a voltage equal to $U_{\mathrm{e}} + m_{\mathrm{d}}U_{\mathrm{e}}$, where m_{d} is the transformation ratio between the first and demagnetization windings. When the demagnetization current reaches zero, then the diode D_{m} turns off, and the voltages U_1 and U_2 are null.

During this last phase, corresponding to the free-wheeling phase of a buck, the diode D_2 is turned on, while the current in L_s decreases.

The voltage U_s applied to the output stage of the converter is then similar to that of a buck converter, when taking into account the transformation ratio m of the transformer. The equations are as a consequence similar: $U_\mathrm{so} = mDU_\mathrm{e}$ where D is the duty cycle. The ratio between the output and the input voltage is a function of D and m (constant). However, the duty cycle cannot take all its theoretical possible values (from 0 to 1) but must be limited because of the demagnetization process. Considering m_d as the transformation ratio between the first and demagnetization windings, D can only vary from 1 to $m_\mathrm{d}/1 + m_\mathrm{d}$.

3 DC/DC converters with an intermediary AC stage

3.1 Generalities

As described in Section 2.4, it is useful for some applications to insert a transformer into the DC/DC converter, to ensure voltage/current adaptation and galvanic isolation. The advantage is that the working frequency of the transformer is defined by the switching frequency of the converter, which is much higher than the frequency of the AC feeding network. This enables a useful reduction of the weight and size of the transformer compared to the conventional solution where the transformer is directly coupled to the AC source.

However, the main drawback of the flyback and the forward converter is that these topologies impose large constraints (over-voltages) on the main active component. Moreover, as switching mechanisms are defined by hard commutation rules, it is difficult to obtain a high efficiency for mid- or high-power applications. In order to solve these difficulties, one solution is to place the middle frequency transformer in an AC stage of the energy conversion chain instead of placing it in a direct DC/DC converter, as illustrated in Fig. 15.

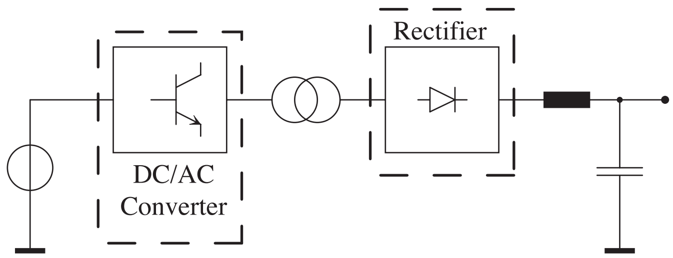


Fig. 15: Main topology

From a voltage source (the capacitor of the input filter in Fig. 3), a first stage consists of a DC/AC converter that feeds the middle frequency converter with an alternative voltage. The DC to AC converter belongs to the family of voltage source inverters.

The secondary side of the transformer is considered as an alternative voltage source. A rectifier is needed to obtain the required DC voltages and currents via an output filter. As we focus our developments on one-quadrant converters, rectifiers are generally non-controlled diode rectifiers. The energy flows are then controlled by the DC/AC converter.

In this conversion chain, the middle frequency transformer cannot be modelled only by its transformation ratio. One must take into account:

– The magnetizing inductor: from the output stage of the DC/AC converter, this inductor ensures that the converter is effectively loaded by the equivalent of a current source (at the commutation process scale).

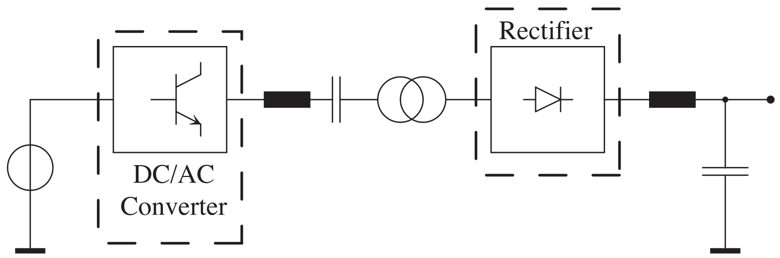


Fig. 16: Main topology: soft commutation

The main difficulty with this inductor is that the DC/AC converter must be controlled to avoid any mean-voltage component that could be applied to the primary side. If this condition is not directly obtained, a blocking capacitor must be inserted in series with the primary side of the transformer, as shown in Fig. 16, to avoid the saturation of the transformer.

- The leakage inductor of the transformer: this inductor has an influence both on the DC/AC converter and on the rectifier, because it limits current variations during each commutation:
  - for the DC/AC converter, this effect can be positive because it allows soft switching conditions that increase efficiency. In some cases, this effect is reinforced by adding a non-linear inductor, series-connected with the primary winding of the transformer as shown in Fig. 16,
  - for the non-controlled rectifier, the limitation of current variations during commutations leads to distortions for the output voltage that affect the averaged value of the regulated output voltage.

When designing a double-stage converter, the magnetizing and the leakage inductors cannot be neglected, and are taken into account as shown in Fig. 17.

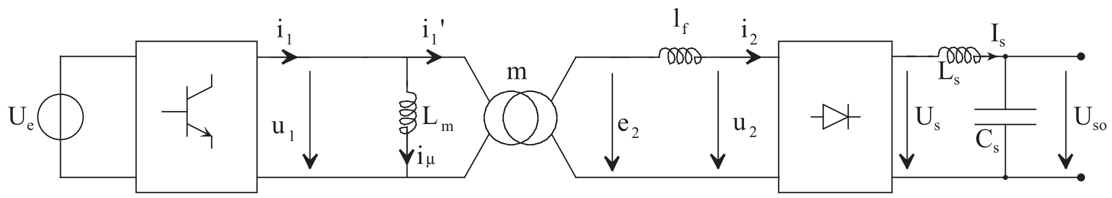


Fig. 17: Integration of the main parameters of a transformer

We shall first describe the main topologies for the DC/AC converter, and describe how such structures offer soft-switching conditions. We then describe the main topologies for the non-controlled rectifier [2].

Finally, we present the working principle of the whole structure, taking into account the model of the transformer and additional component to allow soft-switching conditions.

3.2 DC/AC converters

3.2.1 Main topologies

The aim of DC/AC converters, also called voltage inverter, is to interface a DC voltage source with an AC current source. In the application we define here, the AC source is the primary side of the transformer. As such a device is characterized by its magnetizing and leakage inductors, it can effectively be considered as an AC current source, with respect to commutation mechanisms.

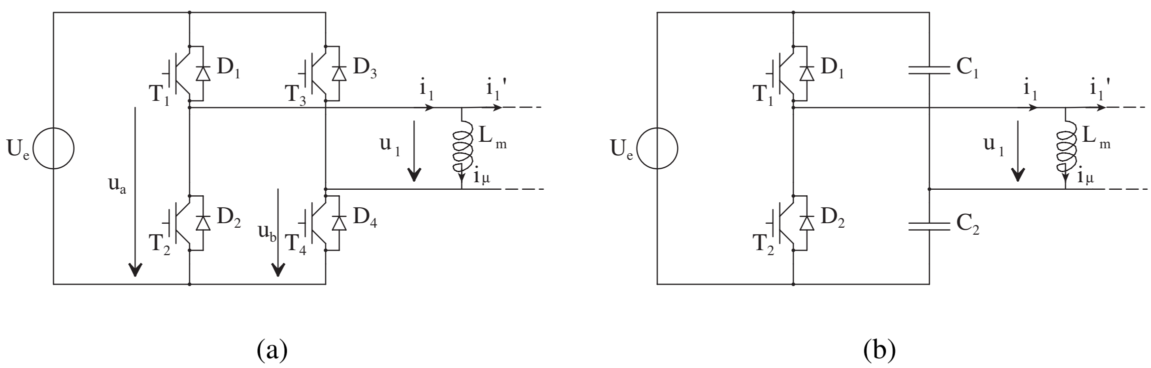


Fig. 18: Voltage inverter: (a) H-bridge. (b) Half-bridge.

The two main topologies that define a voltage inverter are presented in Fig. 18.

Whatever the structure, each switch must be bi-directional for the current, as the current source is an AC source. This is why each transistor is associated with an anti-parallel diode:

- Fig. 18(a): this structure is made of four interrupts that constitute two commutation cells:
  - the first cell is made of (T_1, D_1) and (T_2, D_2),
  - the second cell is made of (T_3, D_3) and (T_4, D_4).

  This structure is interesting because it enables the feeding of the primary side of the transformer with three different voltage levels that can be directly obtained by the controls of the converter:
  - $u_1 = U_e$ when (T_1, D_1) and (T_4, D_4) are turned on,
  - $u_1 = -U_e$ when (T_2, D_2) and (T_3, D_3) are turned on,
  - $u_1 = 0$ when (T_1, D_1) and (T_3, D_3), or (T_2, D_2) and (T_4, D_4) are turned on.
- Fig. 18(b): this structure is obtained from the H-bridge by replacing the second cell [made of (T_3, D_3) and (T_4, D_4)] by two capacitors C_1 and C_2. Both capacitors must have the same capacitance, in order to share identically the main voltage U_e. The voltage across each capacitor is then $U_\mathrm{e}/2$, and must be maintained at this value. This can be achieved by the control of the converter, where conduction time of (T_1, D_1) and (T_2, D_2) must be equal. If this is not the case, sharing resistors must be connected in parallel with each capacitor.
  Two different voltages levels can be directly obtained by the control of the converter:
  - $u_1 = U_\mathrm{e}/2$ when (T_1, D_1) are turned on,
  - $u_1 = -U_\mathrm{e}/2$ when (T_2, D_2) are turned on.

  The level $u_1 = 0$ is also possible, but cannot be forced by the control of the converter. To obtain this level, two conditions must be satisfied:
  - both T_1 and T_2 must be turned off,
  - no current exists at the primary side of the transformer.

  If this last condition is not satisfied, then the diodes D_1 or D_2 will be on, and will be turned off by a decrease to 0 of the current i_s. This depends on the load of the transformer, and not on the control of the DC/AC converter.

As the H-bridge [Fig. 18(a)] offers more flexibility in the voltage levels that are possible for u_1, and as these levels are twice those for the half-bridge [Fig. 18(b)] ($\pm U_\mathrm{e}$ against $\pm U_\mathrm{e}/2$), the H-bridge is most commonly used.

Two different control modes are possible for this type of converter: the rectangular mode and the Pulse Width Modulation (PWM) mode. In one-quadrant DC/DC conversion applications, the rectangular mode is well dedicated, and consists of the control of each commutation cell with a 50% duty cycle, adjusting the energy transfer by a regulation of a phase shift between the two cells, as shown in Fig. 19.

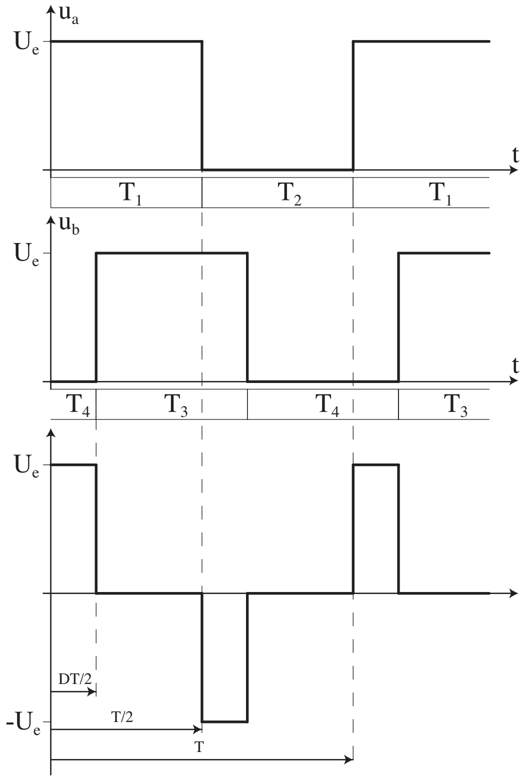


Fig. 19: Rectangular modulation

For the first commutation cell, T_1 and T_2 are alternatively turned on and off with a constant period T and a 50% duty cycle. During the first half of the switching period, the voltage u_a of Fig. 18(a) is U_e, and then 0 during the second half of the period.

The second commutation cell is controlled the same way, except that it is phase-shifted compared to the first commutation cell. This phase shift is defined by the duty cycle D, and refers to half of the switching period ($T/2$).

Then, the resulting voltage u_1 applied to the primary side of the transformer is the difference between u_a and u_b. It consists of positive and negative pulses, with a magnitude equal to $\pm U_\mathrm{e}$, and a width defined by the duty cycle D. It allows the management of the r.m.s. value of the voltage u_1.

3.2.2 Hard commutation – soft commutation

To design power converters with high efficiency, one must analyse switching processes in the commutation cell of the DC/AC converter, in order to establish rules to lower commutation losses. For this, we refer to the typical commutation cell (Fig. 20) of any voltage inverter.

The switching on and off of the diodes is systematically operated with a zero voltage and a zero current, respectively, thus defining spontaneous commutations.

For transistors, one must check the voltage across them and the current through them before turning them on. Various cases must be considered are summarized in Fig. 21.

Figure 21(a) and Fig. 21(b) are related to cases where the period of the AC current source (the primary side of the transformer) is much higher than the switching period of the commutation cell.

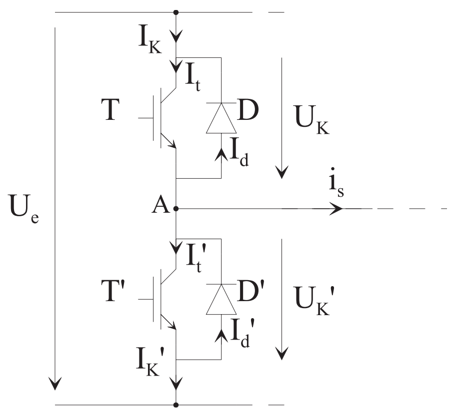


Fig. 20: Commutation cell

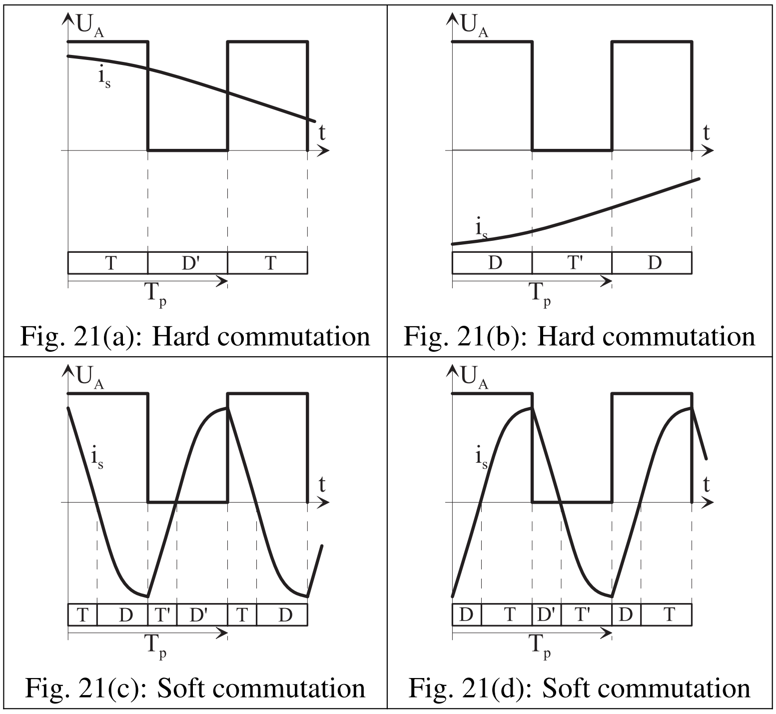


Fig. 21: Types of commutation

One can consider that the commuted current has as a constant sign the switching period scale. Then commutation processes are from one transistor and its opposite diode (T and D', or T' and D). In both cases:

- each switching-on of the transistor is made while the voltage across it is non-zero,
- each switching-off of the transistor is made while the current through it is non-zero,
- each switching-off of the diode is made with reverse recovering current phenomena.

These two switching processes are called hard commutations, because of losses into the transistor and its opposite diode during each commutation.

Figure 21(c) and Fig. 21(d) are opposite cases, where the period of the AC current source (the primary side of the transformer) and the switching period of the commutation cell are identical. Depending on the phase shift between the AC current and the voltage U_{A}, one can obtain

- Fig. 21(c), the phase shift is negative: each turn-off of the transistor is made while the current through it is null. This will lower losses compared to a pure hard commutation process. Such transitions are called Zero-Current Switching (ZCS).
- Fig. 21(c), the phase shift is positive: each turn-on of the transistor is made while the voltage across it is null. This lowers losses compared to a pure hard commutation process. Such transitions are called Zero-Voltage Transition (ZVT).

Because of their reduced commutation losses, ZCS and ZVT commutations are called soft commutations. The main difficulty is that the conditions that do or do not allow soft commutations are directly linked to the commuted current, that means linked to the current at the primary side of the transformer. To ensure that the voltage inverter stage of a one-quadrant DC/DC converter is able to work in soft commutation conditions, one must have already defined the rectifier and output stage that will define the current circulating in the primary side of the transformer during commutations of the DC/AC converter.

3.3 AC/DC converters

These structures are conventional in power electronics. These converters are power interfaces between an AC voltage source (the secondary winding of the transformer) and a DC current source (the inductor of the LC output filter). The generic structure that matches this requirement is given in Fig. 22(a).

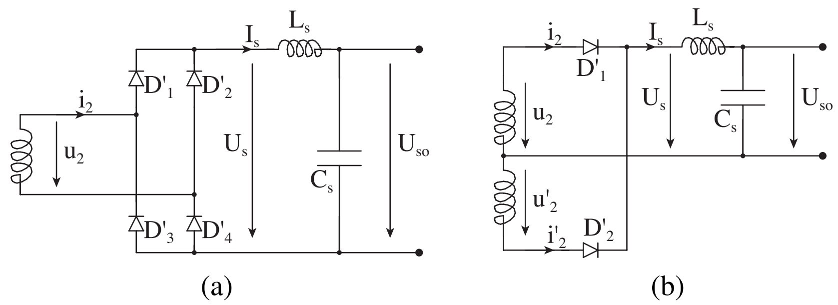


Fig. 22: Non-controlled rectifier: (a) Full bridge. (b) Transformer with two secondary windings.

The use of four diodes induces the current non-reversibility of this converter. Moreover, the generated voltage U_{s} can never be negative:

- when the AC voltage u_2 is positive, then D'_1 and D'_4 are in their on-state, and $U_{\mathrm{s}} = u_2$, $i_2 = I_{\mathrm{s}}$;
- when the AC voltage u_2 is negative, then D'_3 and D'_2 are in their on-state, and $U_{\mathrm{s}} = -u_2$, $i_2 = -I_{\mathrm{s}}$.

The sign of u_2 defines the state of the diodes (on or off), and as a consequence the value of the AC current i_2 is also a function of the output current I_{s}.

Another possibility is to use a two-secondary-windings transformer. The main advantage here is that this leads to a reduction of the diode number, as shown in Fig. 22(b). The voltages u_2 and u'_2 are identical, defined by $u_2 = u'_2 = mu_1$ where u_1 is the voltage on the primary side of the transformer and m the transformation ratio between one secondary winding and the primary one. The working principle is identical to that in the previous structure:

- when the AC voltages u_2 and u'_2 are positive, then D'_1 is in its on-state, and $U_{\mathrm{s}} = u_2$, $i_2 = I_{\mathrm{s}}$;
- when the AC voltages u_2 and u'_2 are negative, then D'_2 is in its on-state, and $U_{\mathrm{s}} = -u_2$, $i'_2 = I_{\mathrm{s}}$.

Because of the reduced number of diodes, we retain this structure for the following section where the full structure of the isolated DC/DC converter is presented, and where its working principle is illustrated.

3.4 Full structure

3.4.1 Ideal transformer

The full structure of a one-quadrant DC/DC converter with an intermediary AC stage is given in Fig. 23, where we consider the transformer as ideal, only defined by its voltage transformation ratio m.

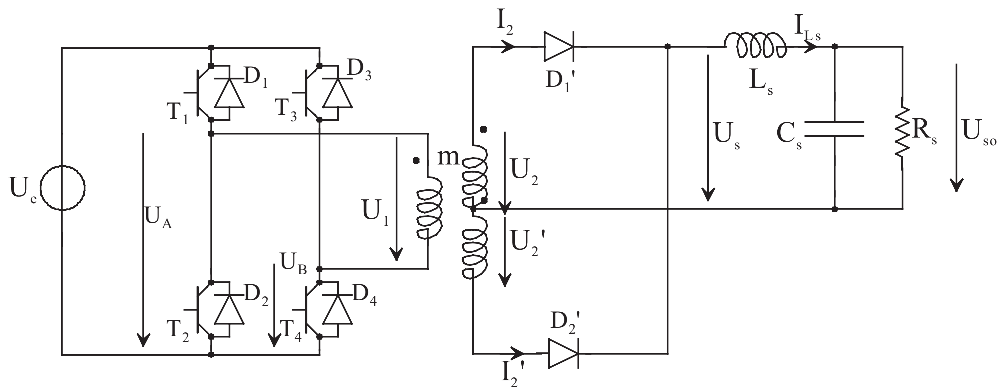


Fig. 23: Full structure with ideal two-secondary-windings transformer

The typical waveforms that illustrate its behaviour, in the case of continuous conduction mode, are given in Fig. 24.

The voltages U_2 and U_2' generated at the secondary side of the transformer are defined by the voltage at the primary side as defined in Fig. 19, and by the transformation ratio m. This voltage is rectified by the non-controlled diode converter, to obtain the output voltage U_s filtered by the $\mathrm{L_sC_s}$ output filter. The voltage U_s is in positive and pulses, with a magnitude equal to mU_e, and a width defined by the duty cycle D. The averaged value of U_s can then be regulated, with the same relation used for the forward converter: $U_\mathrm{so} = mDU_\mathrm{e}$.

On the primary side of the transformer, the current I_1 is equal to mI_2 when the output diode D_1' is on, and equal to $-mI_2'$ when the diode D_2' is on. This current is then effectively an alternative current, with the same period as the switching period of the voltage inverter.

Moreover, particular attention must be given to the conduction sequence of the interrupts of the voltage source inverter defined at the bottom of Fig. 24:

- The first commutation cell made of (T_1, D_1) and (T_2, D_2) defines the change of sign for the voltage u_1, and the commutations into the rectifier stage. Then the primary current I_1 is in phase with the voltage U_A generated by this commutation cell, and the diodes are not solicited.
- The second commutation cell made of (T_3, D_3) and (T_4, D_4) defines the width of output voltage pulses, and of course the duration of the free-wheeling mode. There, the primary current I_1 is not in phase with the voltage U_B generated by this commutation cell. This means that the sign of this current changes while one interrupt is on $[(T_3, D_3)$ or $(T_4, D_4)]$. As a consequence, the diodes of this cell are solicited, so that ZVT commutations directly exist thanks to the free-wheeling mode of the output stage: each switching-on of T_3 and T_4 is made while the voltage across them is null.

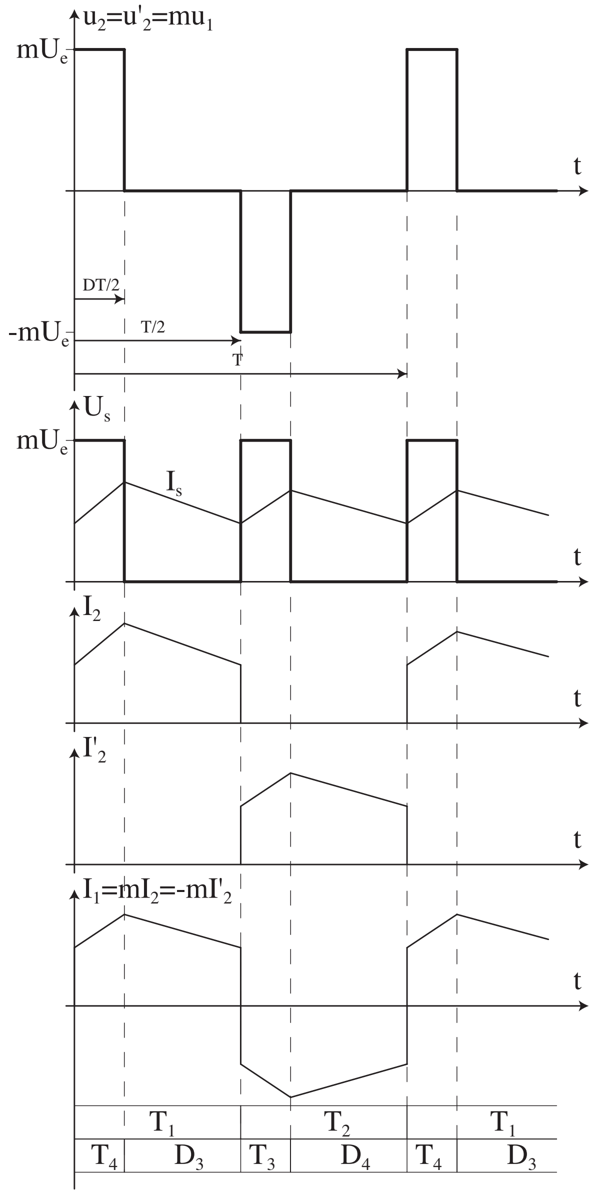


Fig. 24: Typical waveforms

3.4.2 Effects of leakage inductors

As we have defined the control of the voltage source inverter in Fig. 19, we can regulate the DC output voltage U_{so} with soft commutations on the second commutation cell and hard commutations on the first one. This is true when the transformer is considered only with its transformation ratio. Things are different when leakage inductors are taken into account, see Fig. 25.

With leakage inductors, the currents through the secondary windings of the transformer cannot vary instantaneously for each commutation in the diode rectifier. The leakage inductors l_2 and l_2' seen at each secondary winding limit the derivative of the currents I_2 and I_2'. As a result, the diodes D_1 and D_2' are simultaneously in their on-state throughout their commutation processes. One can define α as the duty cycle that defines the duration of this phenomenon, which is defined by the equation $\alpha = \frac{2l}{T}\frac{I_{\mathrm{s}}}{mU_{\mathrm{e}}}$ (l is the leakage inductor seen on one secondary winding, T is the switching period of the voltage inverter, and m the transformation).

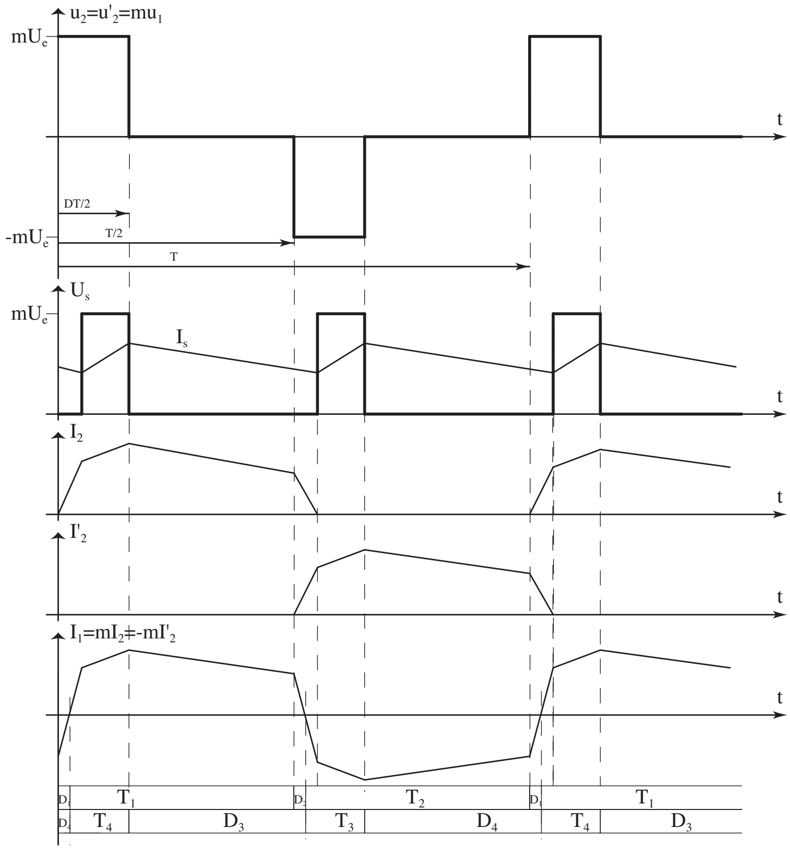


Fig. 25: Typical waveforms

The output voltage is then necessarily affected by the voltage drop caused by the simultaneous conduction of the diodes, and depends not only on the duty cycle D, but also on the output current, $U_{so} = mDU_e - \frac{2l}{T} I_s$, that the regulator of the converter must compensate. This is not necessarily a positive effect.

However, a positive effect can be seen in the conduction sequence of the switches of the voltage inverter. Indeed, if nothing has changed for the second commutation cell [(T_3, D_3) and (T_4, D_4)], then attention must be given to the first commutation cell, made of (T_1, D_1) and (T_2, D_2), that defines the change of sign for the voltage u_1, and commutations into the diode rectifier. During these last commutations, as the current in the secondary windings cannot change instantaneously, the variations of the current at the primary side are also finite. The interrupts of the first commutation cell are solicited, and each effective turn-on of the transistor is made while its associated diode is conducting. As a main result, the first commutation cell works in ZVT soft-commutation conditions, owing to the leakage inductor of the transformer.

3.4.3 Soft-switching configuration

The leakage inductor of the transformer allows ZVT soft-switching conditions for the whole voltage inverter stage. One direct use of this property could be to design the transformer with given values for leakage inductors, allowing soft-switching conditions for a large range of operating points. However, the main drawback is that the averaged output voltage ought to be strongly affected by this, especially for high DC current. This is why additional devices can be inserted into the conversion chain to assure soft-switching conditions in a large range of operating points, independently of the values of the leakage inductors of the transformer.

As an example, one can cite the addition of a blocking capacitor C_b and saturable inductor L_{sat}, series-connected to the primary side of the transformer (Fig. 26).

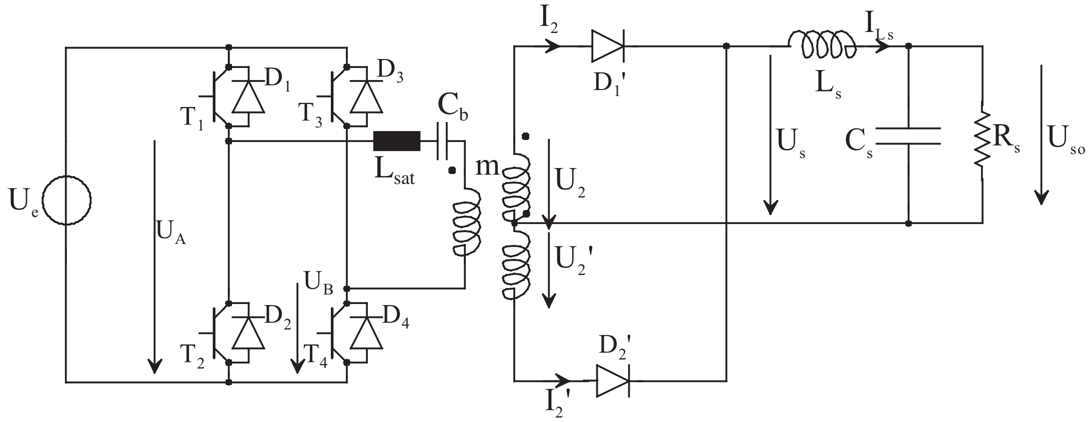


Fig. 26: Additional devices for soft-switching conditions

The blocking capacitor has a double role. The first one is to allow the extinction of the transformer primary current with a controlled derivative over time, at each turn-off of T_1 and T_2. Secondly, it avoids the feeding of the transformer with non-null averaged voltage, that could induce the saturation of the transformer.

The saturable inductor is needed when the commands on T_1 and T_2 are no longer complementary commands, where a dead-time is defined to allow the transformer primary current to stay null during t_d, as shown in Fig. 27.

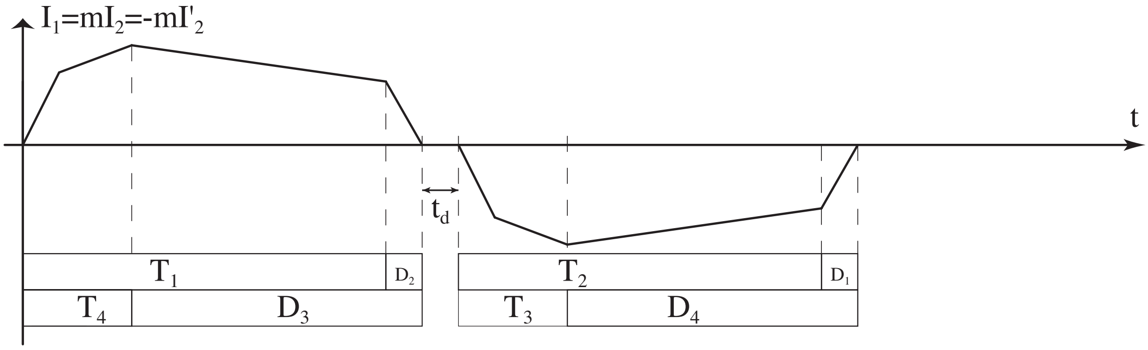


Fig. 27: Primary current waveform

With such a control, both commutation cells are working under ZCS soft-switching conditions. The second one is also naturally characterized by ZVT soft-switching conditions. Depending on the control of the voltage source inverter, this can be modified compared to what is proposed in Fig. 19; it is possible to obtain ZCS and ZVT soft-commutation conditions for the whole DC/AC converter [6].

The solution defined in Fig. 26 can also be adapted to offer soft-commutation conditions for the voltage inverter, by replacing the blocking capacitor by clamping diodes [7], or by adding a resonant active sub-converter in parallel with the voltage source U_e [8], or in parallel with the switches of the voltage inverter [9].

4 Conclusion

We have presented various solutions for realizing one-quadrant, switched-mode power converters. These solutions have been grouped into three families of converters:

– Direct DC/DC converters, generally dedicated for low- to mid-power applications. Their main drawback is that commutation losses can be significant because of hard commutation processes.
– Direct DC/DC converters with transformers (flyback and forward), where the same comments can be made as for the previous family.
– DC/DC converters with intermediary AC stage. These converters can be naturally able to offer soft-commutation conditions that can be reinforced by the insertion of additional LC components or active devices. The gain here is that such structures offer high efficiency on account of low commutation losses. This enables high switching frequency, and a reduction of the size and weight of applications, even for high-power applications.

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