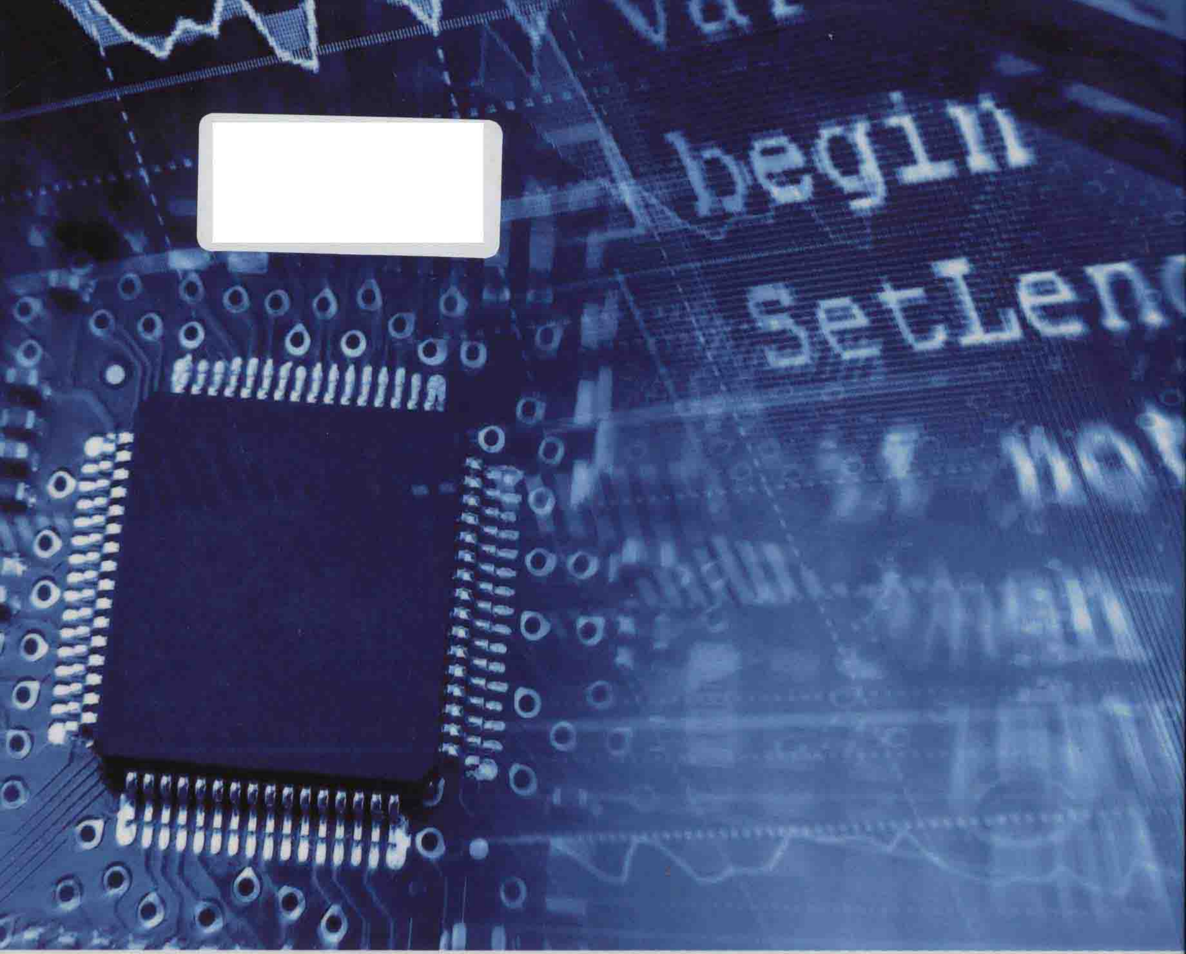

Power Electronics Applied to Industrial Systems and Transports

Nicolas Patin

volume 3

Switching Power Supplies

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Power Electronics Applied to Industrial
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Preface

Volume 3 of this series deals with a specific category of converters for power electronics in the form of switch-mode power supplies. The main function of these components is to provide a continuous voltage to a load, smoothed by filtering elements (and sometimes regulated). In many cases, this requires a high-quality voltage supply (e.g. for electronic chips including components such as microprocessors) to guarantee successful operation. In this context, linear power supplies (ballast type) are often used, but switch-mode supplies are increasingly widespread, allowing high efficiency, and potentially improving the battery life of mobile equipment, for instance, alongside a reduction in the size of cooling elements and/or component heating.

Two main families of switch-mode power supplies will be considered in this volume: non-isolated power supplies (*buck*, *boost* and *buck–boost*) will be covered in Chapter 1, while Chapter 2 will cover isolated power supplies (*flyback* and *forward*). The list of structures presented does not provide exhaustive coverage of topologies found in publications on the subject, but it covers most requirements and includes most of the solutions used in industrial contexts. Note, however, that all of these converters operate using "hard" switching (i.e. with significant switching losses). This means that frequency increases are difficult, or impossible, preventing the miniaturization of passive components used in filtering. To overcome this difficulty, some converter topologies use the resonant behavior of "LC"-type cells, generating a considerable reduction in losses by carrying out "soft" zero voltage switching (ZVS) or zero current switching (ZCS). These converters will be presented in Chapter 3, based on a structure using a resonant inverter associated with a rectifier. The regulation issue mentioned above belongs to the field of automatics (and lies outside the scope of this book), but proportional integral (PI) regulator tuning approaches (for closed-loop control) generally pass through a modeling stage. Chapter 4 presents Middlebrook's approach for average modeling of switch-mode power supplies, which is used to define converter transfer functions. This method will be applied to non-isolated power supplies in order to establish models for continuous conduction mode. We will simply describe the models used in discontinuous mode, which are more difficult to obtain. In conclusion to this volume, we will present a case study of the detailed design and dimensioning of a *flyback* power supply, including the choice of power components (e.g. transistor, diodes, coupled inductances and capacitors) and control elements (e.g. metal oxide semiconductor field effect transistor (MOSFET) gate driver, isolated voltage and current measurements) connected to a microcontroller.

This volume also includes two appendices. Appendix 1 provides general formulas for electrical engineering (and is identical to that included in previous volumes). Appendix 2 supplies the full data sheets for key components of the flyback power supply studied in Chapter 5.

Nicolas PATIN
Compiègne, France
February 2015

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1

Non-Isolated Switch-Mode Power Supplies

1.1. Buck converters

The buck converter is a single-quadrant chopper, as studied in Chapter 1 of Volume 2 [PAT 15b]. The "load" is made up of an inductance L in series with the association of the actual load (presumed to be a current source I_s) in parallel with a filtering capacitor C (see Figure 1.1).

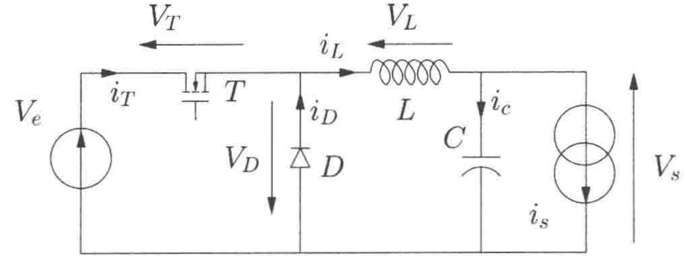


Figure 1.1. *Diagram of a buck converter*

In these conditions, for a correctly dimensioned power supply, the assembly (I_s, C) may be considered to be analogous to the electromotive force (e.m.f.) E_a of a direct current (DC) machine, and the inductance L may be considered to play the same role as the armature inductance in the machine. Consequently, the results established in

Chapter 1 of Volume 2 [PAT 15b] are applicable here. In the case of continuous conduction, an output voltage of $V_s = \alpha.V_e$ is obtained, where α is the duty ratio of the transistor control. Moreover, in cases of discontinuous conduction (i.e. for a current which cancels out in the inductance), the output voltage will be higher than in the continuous conduction case, in accordance with the characteristic shown in Figure 1.4 (see Chapter 1 of Volume 2 [PAT 15b]). A summary of the characteristics of this converter is shown in 1.1. The constraints applicable to the switches are similar to those for a one-quadrant chopper powering a DC machine, but we should also analyze the quality of the voltage supplied to the load. The waveforms produced are the same as those shown in Figure 1.2 (Volume 2, Chapter 1 [PAT 15b]), as the ripple of the output voltage V_s is considered to be a second-order phenomenon, negligible when calculating the ripple of current i_L in inductance L (constant V_s, as for the e.m.f. E_a of a DC machine). Thus, this current may be considered (as in the case of a machine power supply) to be a time-continuous, piecewise-affine function, which may be written (presuming that the load current i_s is constant and equal to I_s) as:

$$i_L\left(t\right) = I_s + \widetilde{i}_L\left(t\right) \qquad [1.1]$$

where $\widetilde{i}_L(t)$ is a signal with an average value of zero, with a "peak-to-peak" ripple Δi_L expressed as:

$$\Delta i_L = \frac{\alpha.\left(1-\alpha\right).V_e}{L.F_d} \qquad [1.2]$$

where $F_d = 1/T_d$ is the switching frequency and α the duty ratio of the control of transistor T.

Second, given the current ripple $\widetilde{i}_L(t)$, the (low) ripple of the output voltage $v_s(t)$ may be deduced, insofar as:

$$v_s\left(t\right) = V_s + \widetilde{v}_s\left(t\right) \qquad [1.3]$$

with:

$$V_s = \alpha . V_e \tag{1.4}$$

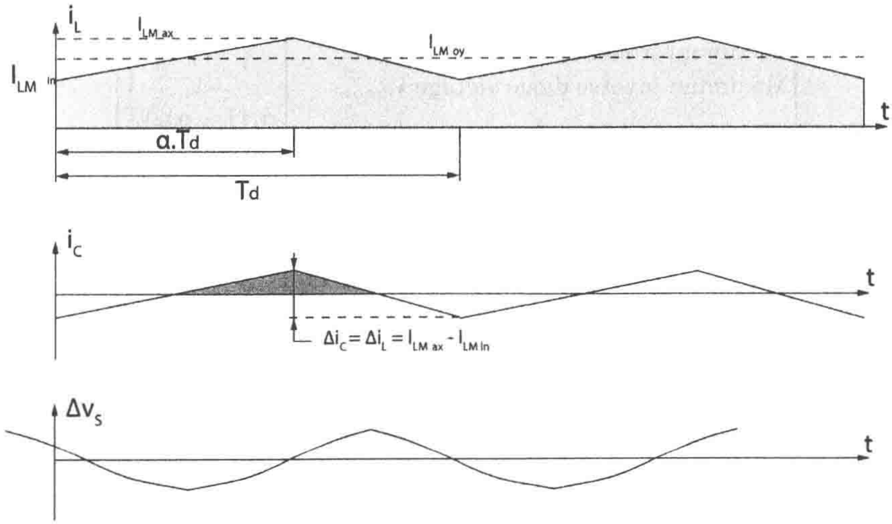
and:

$$\widetilde{v}_s(t) = \widetilde{v}_s(t_0) + \frac{1}{C}\int_{t_0}^{t_0+t} i_C(\tau).d\tau = \widetilde{v}_s(t_0) + \frac{1}{C}\int_{t_0}^{t_0+t} \widetilde{i}_L(\tau).d\tau \tag{1.5}$$

Using this result, it is then easy to deduce the peak-to-peak ripple Δv_s of voltage $v_s(t)$:

$$\Delta v_s = \frac{1}{C.} \cdot \frac{1}{2} \cdot \frac{T_d}{2} \cdot \frac{\Delta i_L}{2} = \frac{\alpha.(1-\alpha).V_e}{8L.C.F_d^2} \tag{1.6}$$

These results are illustrated in Figure 1.2.



Piecewise parabolic evolution of the output voltage

Figure 1.2. *Current in the inductance and voltage at the capacitor terminals in a buck converter*

REMARK 1.1.– In practice, it is important to dimension the capacitor correctly so that the ripple of the output voltage is low in relation to its average value (e.g. 1%) in order to guarantee the validity of the reasoning used above. The current ripple was calculated based on the assumption that the output voltage is constant; strictly speaking, this assumption is not fulfilled, but the simplification is verified in practice. The decoupling of inductance and capacitor dimensioning is widespread when designing switch-mode power supplies, and will be used again when studying other structures. While this reasoning approach may appear artificial, it is based on "auto-coherence" between the initial hypotheses and the desired dimensioning objective. In practice, the output voltage should be as constant as possible when powering electrical equipment using switch-mode supplies (DC/DC converters).

Quantities	*Values*
Maximum transistor voltage V_{Tmax}	V_e
Maximum inverse diode voltage V_{dmax}	$-V_e$
Current ripple in inductance Δi_L	$\dfrac{\alpha.(1-\alpha).V_e}{L.F_d}$
Average output voltage $\langle V_s \rangle$	$\alpha.V_e$
Output voltage ripple Δv_s	$\dfrac{\alpha.(1-\alpha).V_e}{8L.C.F_d^2}$
Maximum current in transistor and diode	$I_s + \dfrac{\Delta I_L}{2}$
Average current in transistor $\langle I_T \rangle$	$\alpha.I_s$
Average current in diode $\langle I_d \rangle$	$(1-\alpha).I_s$
RMS current in transistor (for $\Delta i_L \ll I_s$)	$\sqrt{\alpha}.I_s$
RMS current in diode (for $\Delta i_L \ll I_s$)	$\sqrt{1-\alpha}.I_s$

Table 1.1. *Summary of continuous conduction in the buck converter*

Note that the calculation of the output voltage ripple (used in dimensioning the filter capacitor C) corresponds to a continuous mode of operation. This is not strictly applicable for discontinuous mode, but this choice presents certain advantages:

– simpler calculations using this operating mode;

– the results obtained allow satisfactory dimensioning of capacitors (on the condition that a minimum safety margin is respected) for all possible cases;

– continuous conduction is often the most critical case with regard to the output ripple, although this is not true for buck converters. The output ripple is proportional to the charge current (for boost and buck–boost converters, described in the following sections), and is higher in continuous mode (discontinuous conduction = low charge).

1.2. Dimensioning a ferrite core inductance

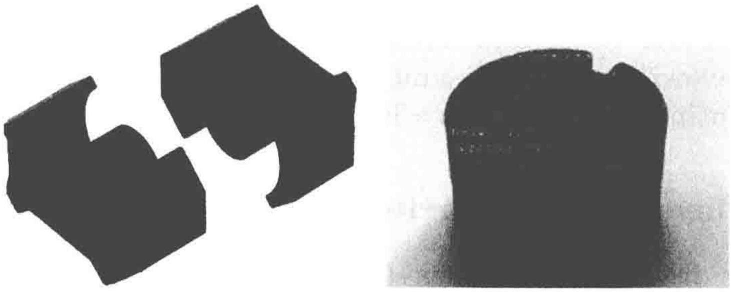
For iron core windings (as discussed in Chapter 5 of Volume 1 [PAT 15a]), a simple magnetic circuit was considered, characterized on the sole basis of three geometric parameters (reduced to two parameters) which needed to be established. While the equation model of the inductance and the applicable usage constraints remain identical, the geometry of a ferrite core is required, and a core must simply be selected from the lists supplied in manufacturer catalogs. The first stage in this process is to choose a family of ferrite cores: this choice depends on the application and the available space. Note the existence of “E,I” structures (alongside double E structures); in power electronics, however, the RM and PM families are most interesting in terms of electromagnetic compatibility, as they are relatively “closed” and produce limited radiation into the immediate environment. Two examples of these families of cores are shown in Figure 1.3.

Once a family of cores has been selected, we need to choose a specific model in accordance with a given specification. To do this, the expressions of A_e (iron section) and S_b (windable section) are used. Note that two surfaces are linked to constraints relating to ferrite (magnetic flux density B_{max}) and copper (current density J_{max}):

$$A_e = \frac{L.I_{max}}{n.B_{max}} \qquad [1.7]$$

and:

$$S_b = \frac{n.I_{RMS}}{K_b.J_{max}} \qquad [1.8]$$



a) b)

Figure 1.3. *RM a) and PM b) ferrite cores*

The obtained product $A_e.S_b$ must be lower than that of the selected core in order to establish an acceptable solution.

Once these geometric elements have been fully designed, we may design an inductance, defining an air gap and calculating the number of turns required in the winding. Note that matching ferrite cores is particularly difficult, and it is therefore best to select a core which already contains an air gap in the central leg. In this case, the manufacturer gives a parameter known as the specific inductance, conventionally denoted as a_L (homogeneous to an inductance). This parameter is simply the inverse of the reluctance of the obtained magnetic circuit. Hence:

$$L = \frac{n^2}{\mathcal{R}_{circ}} = a_L.n^2 \qquad [1.9]$$

in order to calculate the number of turns required.