# Chapter 2
# Introduction to Power Electronic Converters Modeling

This chapter deals with a brief presentation of the main modeling aspects of power electronic converters. The chapter overviews modeling basics, provides useful hints about the main modeling methodologies, accompanied by some illustrative examples, and suggests some possible uses of models.

## 2.1 Models

### *2.1.1 What Is a Model?*

Modeling of a phenomenon or process is based on its observation and relies upon capturing into an approximate, but sufficiently comprehensive, representation, its most significant features from the point of view of a given application. Modeling requires generalization in the sense that the studied phenomenon must be regarded in the context of similar phenomena so common features may be extracted.

Generally speaking, there are two main modeling approaches: one that uses *black-box models*, based on the process behavior observation of its response to some known input signals, and one based on the known information about the system to be modeled (i.e., representation centered on the behavior laws). The latter approach is employed not only to model physical processes, but also biological, economics or even social systems. Mixing between the two approaches is also encountered, leading to the so-called *gray-box models*.

The interest of this textbook is on power electronic converter modeling using the "information" approach. This means that model representations will be made using the available physical knowledge about the considered converter. In general, physical knowledge about system results in mathematical description of mass and energy conservation laws. Thus, energy accumulation variations within the system are described by so-called *state variables*. In the particular case of power converters, information is embodied in Kirchhoff's laws of the converter circuit, Ohm's laws for the various loads and, finally, in the states of various solid-state switches.

S. Bacha et al., *Power Electronic Converters Modeling and Control: with Case Studies*, Advanced Textbooks in Control and Signal Processing, DOI 10.1007/978-1-4471-5478-5_2,

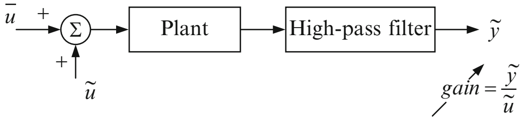



**Fig. 2.1** Basic idea of linear identification approach, where $\bar{u}$ and $\tilde{u}$ are input's low-frequency and high-frequency components, respectively

## 2.1.2 Scope of Modeling

Next within this textbook one seeks to obtain quasi-general power electronic converter dynamical models to simulate converter dynamic behavior and to construct various control laws. Steady-state converter behavior (static models) can also be obtained, either by zeroing the time derivatives in the dynamical models in the case of DC variables, or by zeroing the derivative of both magnitude and phase in the case of AC variables.

Concerning simulation, a plethora of software renders power converter time-domain behavior in a very precise and reliable way (see, for example, SPICE®, SABER®, MATLAB®). With these programs, however, simulation results are not general. For instance, even if they provide various time waveforms of the internal variables, they do not give direct information about the converter modes. As a consequence, obtaining a model necessary for control purposes cannot be done by using these software packages, at least not directly. Certainly, it is possible to identify a power electronic converter based on the input-output variables evolution file obtained by simulation and to elaborate a frequency-domain model. But, since the quasi-totality of power electronic converters is naturally a nonlinear or linear time-varying system and any linear input-output model depends on the operating point, the information acquired has limited validity.

Figure 2.1 shows how the linearized system looks. Another important drawback of the linear identification approach is that system identification in the frequency domain will be limited at half the switching frequency, according to Shannon's theorem.

An analytical model based on knowledge of the circuit's physical behavior is needed for control purposes. According to the intended use, various levels of modeling can be considered. Model choice also relies on the following criteria:

- the required dynamic or steady-state accuracy;
- whether internal, input or output variables should explicitly appear in the model;
- an acceptable level of complexity;
- the domain of definition.

These requirements are not totally synergistic and often are antagonistic, necessitating an optimal choice. For example, the accuracy of plant replication increases with model complexity.

## 2.2 Model Types

One can specify some simplifying assumptions, sufficiently accurate so as not to affect the validity of the models to be implemented:

1. Switches are considered 'perfect' in the sense that they behave as a zero-value resistance during conduction (the so-called ON state) and as an infinite-value resistance when the switch is turned off (the so-called OFF state). Also, the switching time is infinitely short.
2. Generators are considered 'perfect' (for example, they provide infinite short-circuit power in the case of voltage sources).
3. Passive elements are considered linear and invariant.

If the two first modeling assumptions are easy to understand, the third deserves more attention. Let us consider as an example a nonlinear inductance whose value depends both on time and on the current $i(t)$ passing through it. Inductance voltage is given by the following equation:

$$v(t) = \frac{d}{dt}\left(L(i,t) \cdot i(t)\right)$$

Developing this equation gives a nontrivial expression:

$$v(t) = \left(\frac{\partial L(i,t)}{\partial t} + \frac{\partial L(i,t)}{\partial i} \cdot \frac{di(t)}{dt}\right) \cdot i(t) + L(i,t) \cdot \frac{di(t)}{dt}. \tag{2.1}$$

Being too complicated, Eq. (2.1) is practically unusable in modeling. Moreover, this complexity is not justified, as the first term is typically not important in most of applications.

These remarks reasonably justify the adoption of the above assumptions, which however do not essentially affect the modeling methodology. Obviously, in order to increase modeling accuracy, one can successively add detail to an initial, simplified model. For example, models of circuit elements can be enriched by taking into account the dissipative elements (internal resistance of a power source, winding resistance of a coil, etc.). Figure 2.2 shows how the model of a diode can be enriched.

Single-pole–single-throw (SPST) switches can be implemented in various power electronic devices (Erikson and Maksimović 2001). Examples of their schematics are represented in Fig. 2.3. The single-quadrant SPST symbols (e.g., diode and transistor) and their ideal characteristics are presented in Fig. 2.3a, b. Two-quadrant voltage-bidirectional SPSTs can be of many types. Their common features – interesting from the modeling and control viewpoint – have been unified into a unique schematic that will be used throughout this book, as exemplified in Fig. 2.3c.

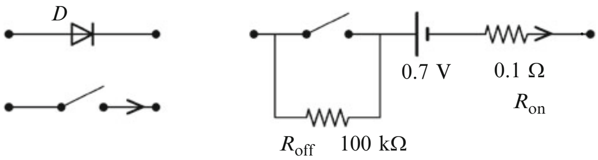



**Fig. 2.2** Diode schematic and ideal model (*left*) and possible enriched model (*right*)

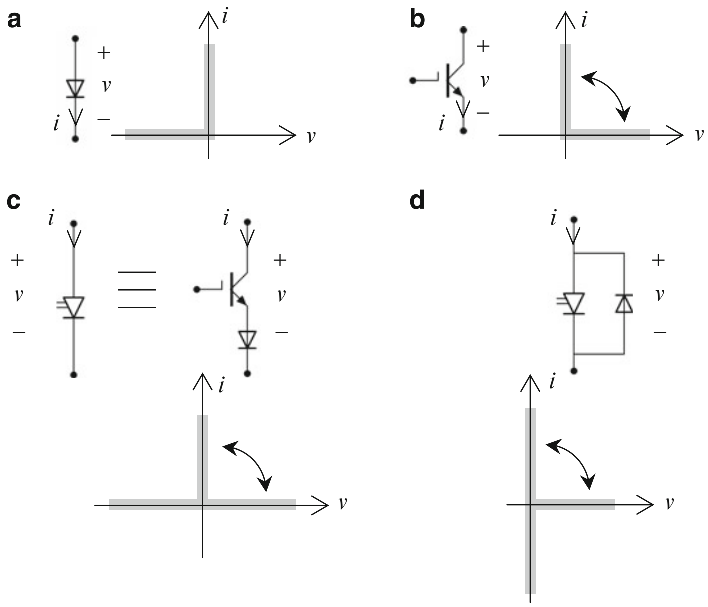



**Fig. 2.3** Various SPST switches and their ideal characteristics: (**a**) diode; (**b**) transistor (BJT/IGBT); (**c**) two-quadrant voltage-bidirectional SPST; (**d**) two-quadrant current-bidirectional SPST (Erikson and Maksimović 2001)

### *2.2.1 Switched Models*

The switched model is the less elaborated model of the converter, meaning it describes the electrical equations for each circuit configuration. It is sometimes named the "*exact*" model because, under the previously stated assumptions, it describes the converter behavior exactly. The reader may refer to Kassakian et al. (1991) or to Erikson and Maksimović (2001) to get an overview of specific converter switched models and their analysis.

Let us consider the example of the buck converter in Fig. 2.4, where the switch is driven by signal $u(t)$, called the *switching function* (Fig. 2.4a). Let us consider that $u(t)$ is periodic, having $T$ as switching period and $\alpha$ as duty ratio:

$$u(t) = \begin{cases} 1, & 0 \le t < \alpha T \\ 0, & \alpha T \le t < T \end{cases}, \quad u(t - T) = u(t) \forall t.$$

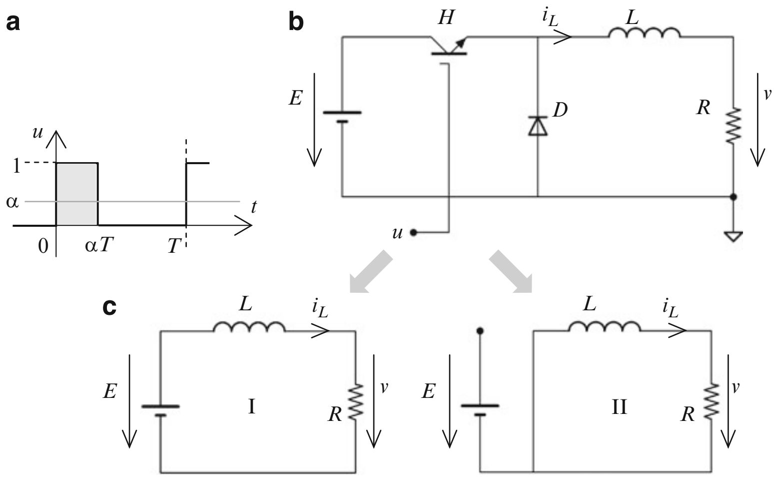



**Fig. 2.4** Example of a buck power stage: (**a**) signal driving switch $H$; (**b**) converter diagram; (**c**) various configurations of the converter: configuration **I** – between time 0 and $\alpha \cdot T$, configuration **II** – between time $\alpha \cdot T$ and $T$

One can easily note that α represents the average value of $u(t)$.

According to the state of the switch $H$, the circuit takes the configuration **I** (switch turned on) and configuration **II** (switch turned off) in Fig. 2.4.

Configuration **I** corresponds to a time $t$ (modulo $T$) between 0 and $\alpha \cdot T$; the system behavior is given by the first equation from (2.2). Configuration **II** corresponds to a time $t$ between $\alpha \cdot T$ and $T$. One may observe that the circuit has in fact two switching devices, as the diode also naturally turns on and off (Sira-Ramírez and Silva-Ortigoza 2006). Hence, its governing equations are:

$$\begin{cases} E = L\dfrac{di_L}{dt} + Ri_L \\ 0 = L\dfrac{di_L}{dt} + Ri_L, \end{cases} \tag{2.2}$$

where $v = R \cdot i_L$. An elegant way to describe this behavior is to use the switching function $u$ and to compress the representation (2.2) into the form below:

$$E \cdot u(t) = L\frac{di_L}{dt} + Ri_L. \tag{2.3}$$

Function $u$ takes the values 1 (switch turned on) and 0 (switch turned off) according to the configuration. This leads to the equivalent electrical diagram presented in Fig. 2.5, which will from now on be called the *exact equivalent circuit*.

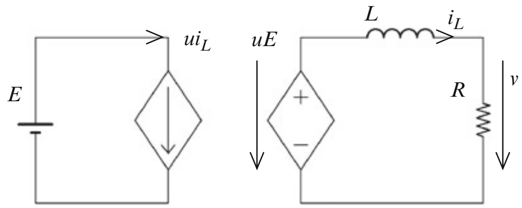



**Fig. 2.5** Exact equivalent circuit of buck power stage showing input port configuration (Erikson and Maksimović 2001)

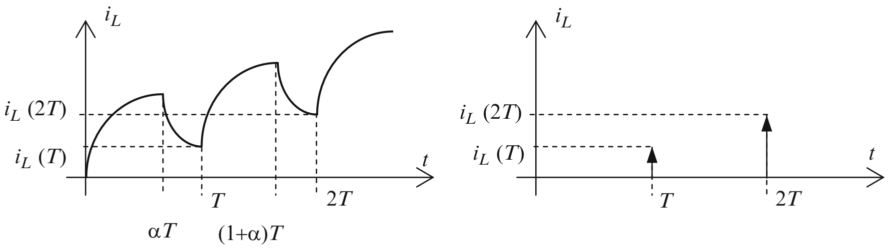



**Fig. 2.6** Buck converter sampling and sampled-data model

### 2.2.2 *Sampled-Data Models*

A sampled-data model is a model that provides information about the system state in a periodic manner. In the present case, it is a representation sampled not at the switching moments but at each complete operating period (Verghese and Stanković 2001). In the buck converter case detailed in Fig. 2.4, the system switches between two circuit configurations and the inductor current time evolution can be like the one presented in Fig. 2.6. If one considers the current values at each switching period $T$, a recurrent equation is obtained as below:

$$i_L((k+1)T) = \left(i_L(kT) - \frac{E}{R}\right) \cdot e^{-\frac{R}{L}T} + \frac{E}{R} \cdot e^{-\frac{R}{L}(1-\alpha)T}. \tag{2.4}$$

Equation (2.4) can be put into the more general matrix form:

$$\mathbf{x}_{k+1} = \Phi(\mathbf{x}_k, \mathbf{u}_k, \mathbf{p}_k), \tag{2.5}$$

where **x**, **u** and **p** are the state, control input and disturbance vectors, respectively. Model (2.5) gives the system state at each sampling period but does not provide any information about the concerned variables between the two sampling points. By virtue of its discrete-time description, this model may be useful in digital control of converters (Maksimović et al. 2001).

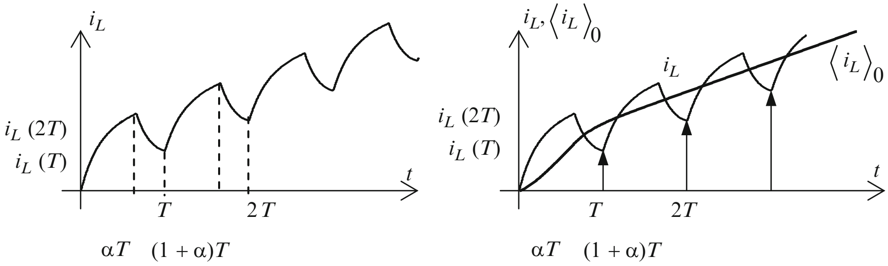



**Fig. 2.7** Behaviors of switched and averaged models

### *2.2.3 Averaged Models*

As their names show, these models replicate an average behavior of the system state. This average does not remain constant for a time period comparable with the system time constant but changes as the system is excited. It is computed on a time window of width $T$, which is sufficiently small in relation to the system dynamics. This window must be regarded as sliding on the time axis, and so it will be called *sliding average* (*moving average* in statistics).

In the case of a chopper inductor current, sliding average is expressed as

$$\langle i_L \rangle_0(t) = \frac{1}{T} \cdot \int_{t-T}^{t} i_L\, d\tau. \tag{2.6}$$

The averaging of the exact model (2.3) gives

$$\frac{d}{dt} \langle i_L \rangle_0 = -\frac{R}{L} \langle i_L \rangle_0 + \alpha \cdot \frac{E}{L}. \tag{2.7}$$

Model (2.7) is the averaged model of the circuit in Fig. 2.4. Its time evolution is shown in Fig. 2.7.

One can remark that the averaged model is less accurate than the sampled-data model at the precise sampling times; conversely, this approach provides information between the sampling points. A thorough analysis of averaged models for specific power stages has been performed by Kislovsky et al. (1991) and by Erikson and Maksimović (2001).

### *2.2.4 Large-Signal and Small-Signal Models*

Converter dynamic behavior is nonlinear with few exceptions. Sometimes, in order to perform a modal analysis or to build linear control laws, it is necessary to develop

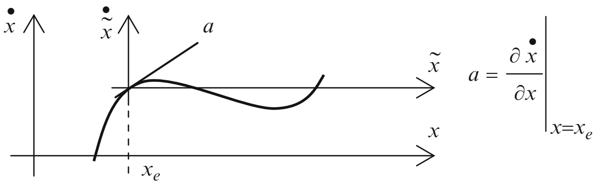



**Fig. 2.8** Relation between large-signal model and small-signal model of a second-order system in the state space

linear models around a certain operating point. To this end the first-order Taylor series expansion is used. These linearized models are valid only for slight variations around the considered operating point. This is why they are called *small-signal models*, also known as *tangent linear models*. Conversely, the initial models, valid on the entire definition range, are called *large-signal models* by the power electronics community.

Figure 2.8 suggests the relation between the large-signal model and the small-signal one in the intuitive case of a second-order system state-space trajectory. If the large-signal model is linear, then it is identical to the small-signal model. This is a quite seldom encountered situation – an example is the ideal buck converter case for constant load value. This approach is similar to both averaged modeling and sampled-data modeling.

Let us consider the general case of a continuous nonlinear system:

$$\begin{cases} \dfrac{d}{dt}\mathbf{x} = f(\mathbf{x}(t), \mathbf{u}(t)) \\ \mathbf{y} = h(\mathbf{x}(t), \mathbf{u}(t)), \end{cases} \tag{2.8}$$

where $\mathbf{x}$, $\mathbf{u}$ and $\mathbf{y}$ are the state, input and output vectors.

#### 2.2.4.1 Obtaining Steady-State Models

By zeroing the derivatives one obtains the steady-state input-output characteristic, i.e., the locus of the system's equilibrium points, denoted with the subscript $e$ and represented by a generally nonlinear curve in the input-output plane:

$$\mathbf{y}_e = g(\mathbf{u}_e).$$

#### 2.2.4.2 Building Small-Signal Models

Let us now consider the small variations $\tilde{\mathbf{x}} = \mathbf{x} - \mathbf{x}_e$, $\tilde{\mathbf{u}} = \mathbf{u} - \mathbf{u}_e$ and $\tilde{\mathbf{y}} = \mathbf{y} - \mathbf{y}_e$ around a given equilibrium point $\mathbf{y}_e$, established in response to input $\mathbf{u}_e$. Thus, the linearized system around the specified equilibrium point is written as

$$\begin{cases} \dot{\tilde{\mathbf{x}}} = \mathbf{A} \cdot \tilde{\mathbf{x}} + \mathbf{B} \cdot \tilde{\mathbf{u}} \\ \tilde{\mathbf{y}} = \mathbf{C} \cdot \tilde{\mathbf{x}} + \mathbf{D} \cdot \tilde{\mathbf{u}}, \end{cases} \tag{2.9}$$

with

$$\begin{cases} \mathbf{A} = \left(\dfrac{\partial f(\mathbf{x},\mathbf{u})}{\partial \mathbf{x}}\right)_{\mathbf{x}_e,\mathbf{u}_e} & \mathbf{B} = \left(\dfrac{\partial f(\mathbf{x},\mathbf{u})}{\partial \mathbf{u}}\right)_{\mathbf{x}_e,\mathbf{u}_e} \\ \mathbf{C} = \left(\dfrac{\partial h(\mathbf{x},\mathbf{u})}{\partial \mathbf{x}}\right)_{\mathbf{x}_e,\mathbf{u}_e} & \mathbf{D} = \left(\dfrac{\partial h(\mathbf{x},\mathbf{u})}{\partial \mathbf{u}}\right)_{\mathbf{x}_e,\mathbf{u}_e} \end{cases}. \tag{2.10}$$

In the case of bilinear systems, i.e., when nonlinearity is given by a product between two state variables or between a state variable and an input variable, another approach is used, as follows.

In the model given by (2.8) one introduces the above defined small variations. Further, some simplifications are made by:

- neglecting the products of variations corresponding to higher-than-second-order terms in the Taylor series expansion;
- simplifying terms corresponding to $\mathbf{x} = 0$.

The resulting model is described by the same matrices as those corresponding to the linearized model (2.9), i.e., those given by (2.10).

*Example*. The example below concerns a bilinear system. The two above-presented approaches will be employed in order to obtain the small-signal model.

$$\begin{cases} \dot{x}_1 = 2x_1x_2 - x_2u \\ \dot{x}_2 = x_1 + x_2 \\ y = x_1^2 + u. \end{cases} \tag{2.11}$$

First, one searches the equilibrium points of the system (2.11) for an input $u = u_e$. This is done by zeroing the $x_1$ and $x_2$ derivatives. By solving the resulting algebraic system two solutions can be found. The first one is trivial, namely $x_{1e} = x_{2e} = 0$, which gives $y_e = u_e$. The other is $x_{1e} = u_e/2$ and $x_{2e} = -u_e/2$, which gives $y_e = 3u_e/4$. Second, the matrices of the linearized model must be obtained. Two methods may be employed as follows:

- *First method*: Using relations (2.10) one obtains matrices **A**, **B**, **C** and **D**:

$$\begin{cases} \mathbf{A} = \begin{bmatrix} \dfrac{\partial f_1(x,u)}{\partial x_1} & \dfrac{\partial f_1(x,u)}{\partial x_2} \\ \dfrac{\partial f_2(x,u)}{\partial x_1} & \dfrac{\partial f_2(x,u)}{\partial x_2} \end{bmatrix}_{x_{1e},x_{2e},u_e} = \begin{bmatrix} -u_e & 0 \\ 1 & 1 \end{bmatrix}, \\ \mathbf{B} = \begin{bmatrix} \dfrac{u_e}{2} & 0 \end{bmatrix}^T, \quad \mathbf{C} = [u_e \quad 0], \quad \mathbf{D} = 1. \end{cases}$$

- *Second method*: One considers the small-signal variations $\widetilde{\mathbf{x}} = \mathbf{x} - \mathbf{x}_e$, where $\mathbf{x}^T = [x_1 \quad x_2]^T$, $\mathbf{x}_e^T = [x_{1e} \quad x_{2e}]^T$, $\widetilde{u} = u - u_e$ and $\widetilde{y} = y - y_e$ which are substituted in equation set (2.11); hence,

$$\begin{cases} \dot{\widetilde{x_1}} + \dot{x_{1e}} = 2(\widetilde{x_1} + x_{1e})(\widetilde{x_2} + x_{2e}) - (\widetilde{x_2} + x_{2e})(\widetilde{u} + u_e) \\ \dot{\widetilde{x_2}} + \dot{x_{2e}} = (\widetilde{x_1} + x_{1e}) + (\widetilde{x_2} + x_{2e}) \\ y + y_e = (\widetilde{x_1} + x_{1e})^2 + (\widetilde{u} + u_e). \end{cases} \tag{2.12}$$

At the steady-state point, the system is described by the relations below:

$$\begin{cases} 0 = 2x_{1e}x_{2e} - x_{2e}u_e \\ 0 = x_{1e} + x_{2e} \\ y_e = x_{1e}^2 + u_e. \end{cases} \tag{2.13}$$

Next, one develops the products. By neglecting the products of variations $\widetilde{x}_1 \cdot \widetilde{x}_2, \widetilde{x_1}^2$ and $\widetilde{x_2} \cdot \widetilde{u}$, knowing that $\dot{x_{1e}} = \dot{x_{2e}} = 0$ and taking into account (2.13), one consequently obtains

$$\begin{cases} \dot{\widetilde{x_1}} = 2x_{1e}\widetilde{x_1} + (2x_{1e} - u_e)\widetilde{x_2} - x_{2e}\widetilde{u} \\ \dot{\widetilde{x_2}} = \widetilde{x_1} + \widetilde{x_2} \\ \widetilde{y} = 2x_{1e}\widetilde{x_1} + \widetilde{u}. \end{cases} \tag{2.14}$$

From (2.13) one computes the steady-state values of the state variables as functions of $u_e$: $x_{1e} = u_e/2$ and $x_{2e} = -\ u_e/2$. By replacing these values, the system (2.14) becomes:

$$\begin{cases} \dot{\widetilde{x_1}} = -u_e\widetilde{x_1} + \dfrac{u_e}{2}\widetilde{u} \\ \dot{\widetilde{x_2}} = \widetilde{x_1} + \widetilde{x_2} \\ \widetilde{y} = u_e\widetilde{x_1} + \widetilde{u}, \end{cases} \tag{2.15}$$

which corresponds to the matrices **A**, **B**, **C** and **D** computed using the first method.

The small-signal model may also be expressed in the frequency domain, i.e., as a transfer function, which can be computed based upon the above deduced matrices:

$$H(s) \triangleq \frac{\widetilde{Y}(s)}{\widetilde{U}(s)} = \mathbf{C}(s\mathbf{I} - \mathbf{A})^{-1}\mathbf{B} + \mathbf{D},$$

where $\widetilde{U}(s)$ and $\widetilde{Y}(s)$ are the Laplace transforms of scalar time signals $\widetilde{u}$ and $\widetilde{y}$, respectively. The same final expression of $H(s)$ can be reached if the Laplace transform is applied to Eq. (2.15). By denoting with $\widetilde{X_1}(s)$ and $\widetilde{X_2}(s)$ the Laplace transforms of the small-signal model state variables, one obtains

$$\frac{\widetilde{X_1}(s)}{\widetilde{U}(s)} = \frac{u_e/2}{s + u_e}, \frac{\widetilde{X_2}(s)}{\widetilde{X_1}(s)} = \frac{1}{s - 1}, \widetilde{Y}(s) = u_e X_1(s) + U(s),$$

allowing us to remark that the state $\widetilde{x_2}$ is unobservable – because it does not appear in the expression of the output – and also unstable.

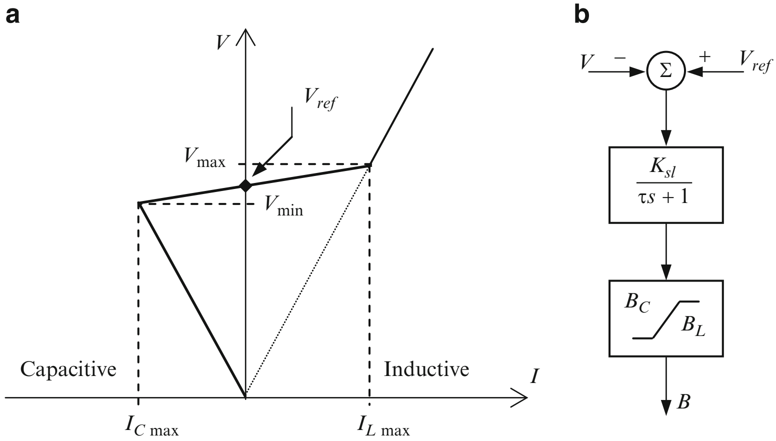



**Fig. 2.9** (**a**) Static regulation characteristic of a SVC (Watanabe et al. 2011); (**b**) elementary dynamic model of a SVC

### 2.2.5 *Behavioral Models*

It is common to use more or less elaborate black-box-type models to replicate the steady-state or dynamic input-output characteristic of *flexible AC transmission systems* (FACTS). Such models are called *behavioral models*.

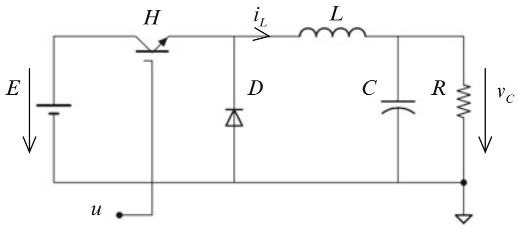
The simplest models are the static ones, as illustrated for example by the case of the *static VAR compensator* (SVC) from Fig. 2.9a. Here one can note a regulation zone described by the following relation:

$$V = V_{ref} + X_{sl}I,$$

where $I$ is the current exchanged by the SVC and the utility grid, $V$ is the voltage at the point of coupling, $V_{ref}$ is the voltage reference value and $X_{sl}$ is the slope of the regulation characteristic. As a matter of fact, FACTS are modeled by means of a reactance or susceptance depending on the desired voltage.

A simple (first-order) dynamical model can be built based upon the static curve from Fig. 2.9a; it can be seen in Fig. 2.9b. The term $K_{sl}$ is the inverse of reactance $X_{sl}$, $\tau$ is the time constant of the system and $B_C$ and $B_L$ are the susceptances corresponding to the SVC's capacitor and inductance, respectively. The model output is the susceptance relative to the error $V_{ref} - V$. Other more or less complicated models are recommended by CIGRE (1995) and IEEE (1993), which rely upon the same principle, matching the input-output characteristics of FACTS.

**Fig. 2.10** Buck converter with second-order filter



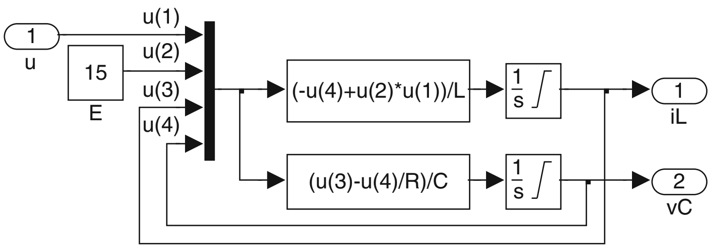




**Fig. 2.11** Example of MATLAB®-Simulink® implementation of exact model of a buck converter

## 2.2.6 *Examples*

Let us consider the case of a buck converter as shown in Fig. 2.10. Via the topological analysis of Fig. 2.4 and Eq. (2.3), the governing equations are

$$\begin{cases} L\dfrac{di_L}{dt} = -Ri_L + Eu(t) \\ C\dfrac{dv_C}{dt} = i_L - \dfrac{v_C}{R}. \end{cases} \tag{2.16}$$

Model (2.16) can express the exact (switched) behavior of the system, as well as its averaged behavior, depending on whether the input signal $u$ is represented by a switched, discontinuous-time-derivative time function or by its continuous-time-derivative average, respectively.

One can attempt a comparison between the different models by means of simulation diagrams; an example of implementation using the Simulink® library is presented in Fig. 2.11. State variables $i_L$ and $v_C$ are the outputs, switching function $u$ and voltage $E$ are the independent inputs. Note also the state variables low limitation required by modeling the unidirectional nature of the converter.

Figure 2.12 allows a comparison between the switched and averaged time evolutions of the two state variables of the system in Fig. 2.11 during the start-up regime, which consists in feeding the system with a duty ratio step. Note the underdamped response due to the unusual choice of $L$ and $C$, the inductor current limitation and the slight difference in capacitor voltage in the two models.

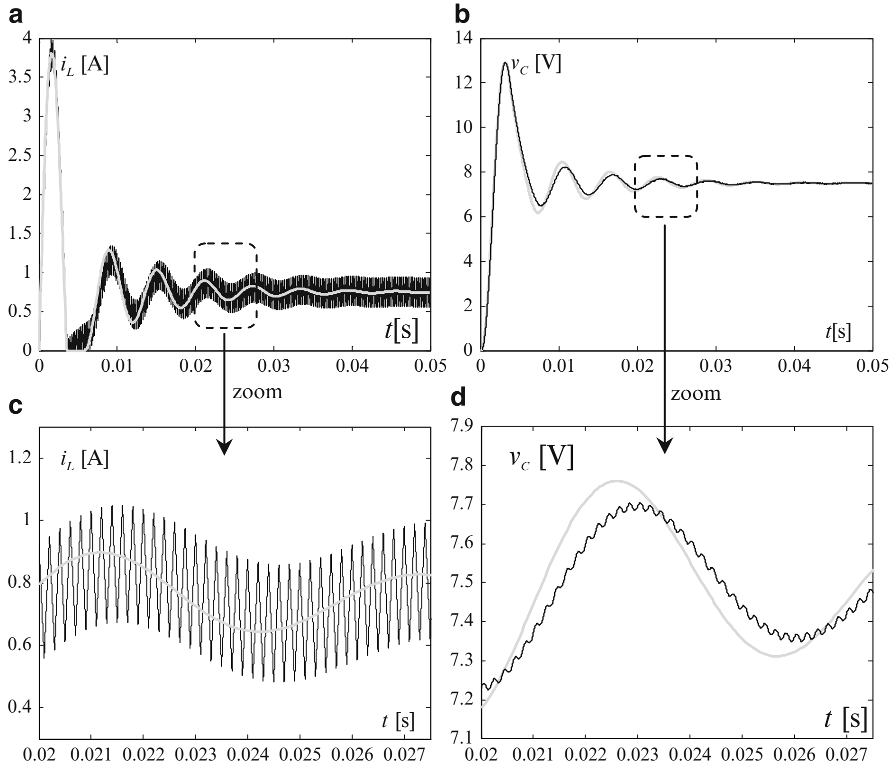



**Fig. 2.12** Switched versus averaged model of a buck converter illustrated by time evolution of state variables at start: (**a**) current $i_L$ evolution, switched (*black*) and averaged (*gray*); (**b**) voltage $v_C$ evolution, switched (*black*) and averaged (*gray*)

As is valid for the entire operating range, model (2.16) is a large-signal one. However, one may be interested in capturing the behavior around a typical steady-state operating point ($u_e$, $i_{Le}$, $v_{Ce}$). In the usual operating range, the form expressed by Eq. (2.14) is linear; this form also characterizes small-signal behavior around the given operating point. The latter belongs to the input-output steady-state model that can be obtained by zeroing derivatives in (2.16):

$$\begin{cases} i_{Le} = \dfrac{E \cdot u_e}{R} \\ v_{Ce} = E \cdot u_e, \end{cases}$$

where the second equation shows the converter's static behavior: the output voltage is obtained by stepping down the input voltage proportionally with the steady-state duty ratio, $u_e$.

Small-signal analysis may result in a transfer-function-type description of the system. In this case, these transfers are defined between the chosen control input (duty ratio) and each of the states.

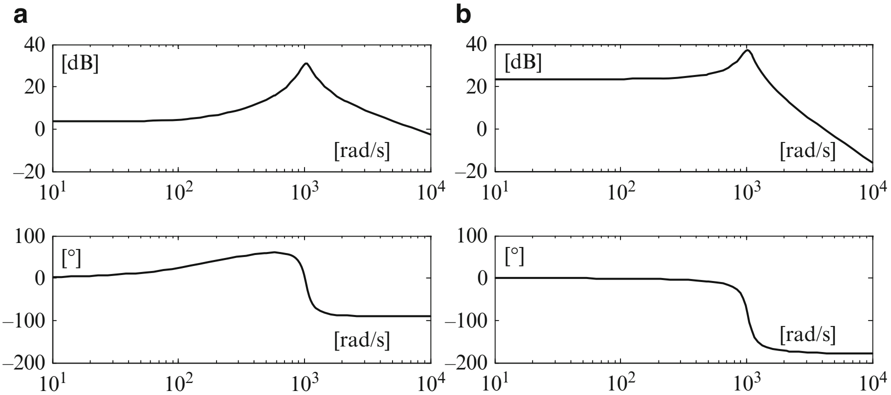



**Fig. 2.13** Small-signal model of a buck converter illustrated in the frequency domain: (**a**) Bode diagrams on channel from input to current $i_L$; (**b**) Bode diagrams on channel from input to voltage $v_C$

Figure 2.13 illustrates a frequency-domain representation of the small-signal, linearized model around the previously stated steady-state operating point. Thus, one can see the Bode diagrams – magnitude and phase – corresponding to the two influence channels from the input represented by the duty ratio to each of the state variables.

The considerably important resonance corresponds to an underdamped time response in Fig. 2.12. These diagrams generally depend on operating point. Moreover, load resistance, which is a parameter in transfer functions, may vary. Therefore, the converter control based on this model should be robust enough to handle all these uncertainties.

## 2.3 Use of Models

### *2.3.1 Relations Between Various Types of Models*

Let us consider a control purpose and a certain control approach to fulfill this purpose.

According to the control law type and to the imposed closed-loop performances, a certain type of modeling is appropriate. Figure 2.14 shows there are two main modeling branches, which lead to discrete-time and continuous-time models, respectively. Many ways of performing conversions between various models are available. One should note that some transformations are more accurate than others.

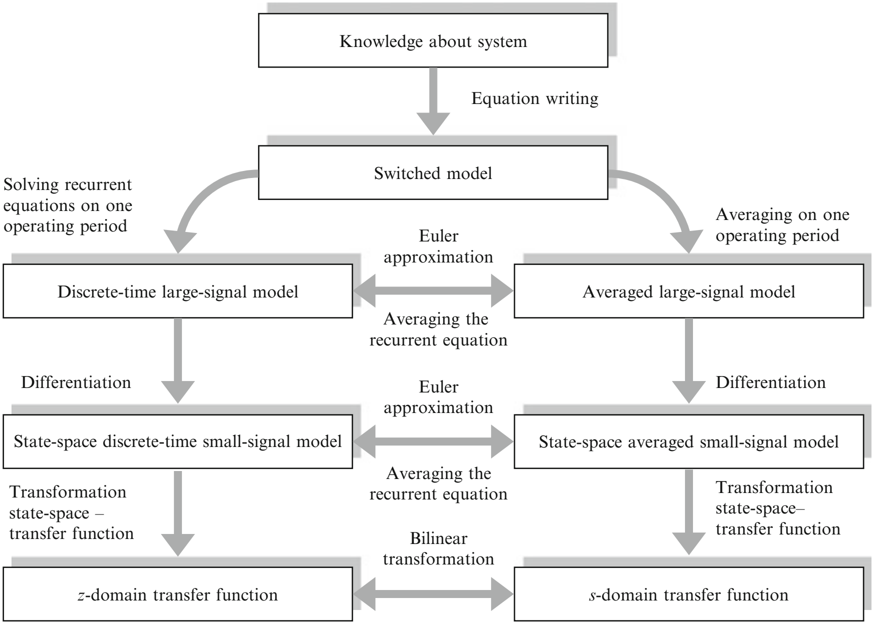



**Fig. 2.14** Relations between different types of models (Bacha and Etxeberria 2006)

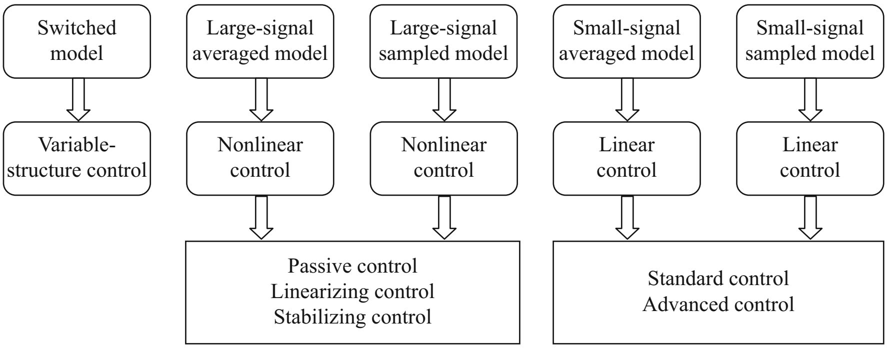



**Fig. 2.15** Relations between models and control laws

## *2.3.2 Relations Between Modeling and Control*

Figure 2.15, showing certain relations between control laws and associated models, is not exhaustive. Indeed, it is possible to build a variable-structure control law using a continuous large-signal model even though this is not the natural approach. However, there are some difficult cases like, for example, tuning a continuous-time proportional–integral controller using a large-signal nonlinear model.

**Table 2.1** Different classes of models used for simulation purposes

| Model | Simulation of dynamic phenomena | Simulation of transient phenomena |
|---|---|---|
| Static (knowledge-based or behavioral models) | Based on modal separation | Not applicable |
| Large- or small-signal averaged models, continuous behavioral models | Depending on the dynamics to emphasize | Emphasizing transients of fundamentals (magnitude and phase) |
| Switched (topological) models | Too computational-time expensive | Emphasizing harmonic phenomena |

### *2.3.3 Other Possible Uses of Models*

Regarding other possible uses of power electronic device models, one can identify roughly three main classes: models used for control purposes, models dedicated to simulation purposes and models employed for sizing and dynamical analysis. If refining simulation purposes further, two classes can be emphasized. Table 2.1 indicates which class of models is suitable for each of the two simulation uses.

Sizing traditionally relies upon static models. Averaged dynamic models are increasingly exploited, especially for emphasizing that certain constraints may be broken. Topological models are widely used to assess the quality of energy, especially those related to harmonic spectra and corresponding filtering.

## 2.4 Conclusion

To conclude, as power electronic converters are, in general, nonlinear variant systems, the modeling operation is a determinant step of every control design attempt. Multiple modeling approaches can be used for this purpose, as this chapter has suggested. A thorough analysis of these models will be provided in the chapters ahead.

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