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# Computation Time Efficient Models of DC-to-DC Converters for Multi-Domain Simulations

Johannes V. Gragger  
 AIT Austrian Institute of Technology GmbH, Mobility Department  
 Austria

## 1. Introduction

In this work power electronic models for DC-to-DC converters, suitable for multi-domain simulations, are presented. Loss calculations in complex electromechanical systems are the main application of the proposed models. Using the proposed power electronic models, the overall efficiency and energy consumption of complex electromechanical systems (e.g. in modern vehicle concepts) working in many different operation points can be simulated with reasonable efforts.

Energy balance is guaranteed and linear temperature dependence of on-resistances, knee voltages and switching losses is considered. Therefore, the modeling approach allows a consistent calculation of the energy flow (electrical, mechanical, thermal, etc.). A detailed outline of the analytical approach used for three DC-to-DC converter models is given. Buck (fig. 1), boost (fig. 2) and buck-boost converter models (fig. 3) with consideration of switching and conduction losses are described.

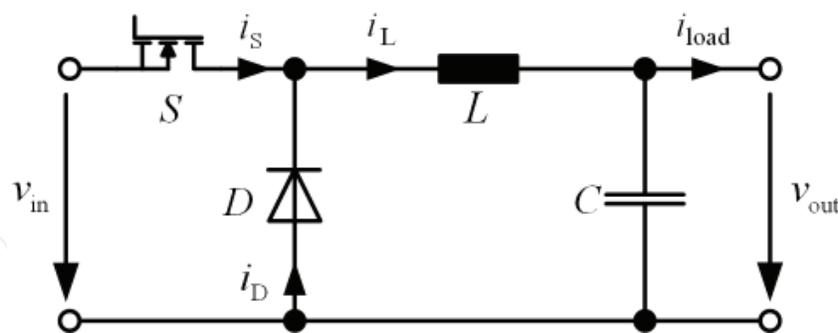


Fig. 1. Topology of a buck converter.

## 2. System design using multi-domain simulation

The development of complex electromechanical systems (e.g. systems in HEVs, EVs, processing plants, power stations, etc.) is very challenging because of the huge number of interacting components. When specifying the core components, the interactions of all components have to be taken into account. These interactions are difficult to overlook if the total number of components is high. In many cases such complex systems are verified through very cost intensive prototyping. For systems in processing plants, etc. the design

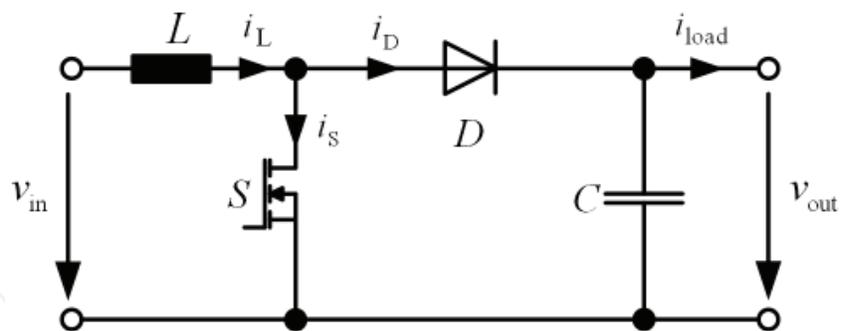


Fig. 2. Topology of a boost converter.

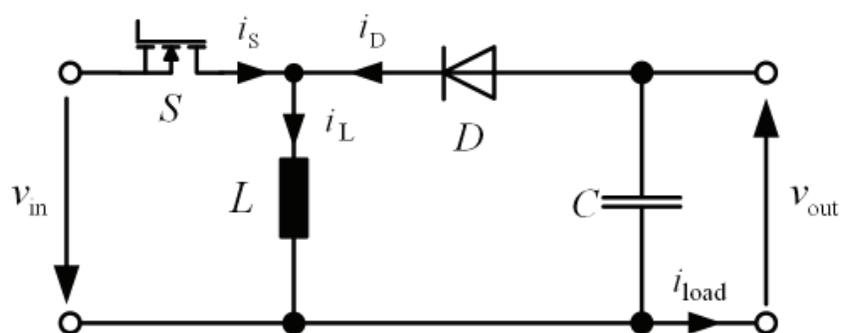


Fig. 3. Topology of a buck-boost converter.

verification is even more difficult. Utilizing computer aided design tools to improve the design process has become state-of-the-art. However, many of the simulation tools that are widely used in industry are specialized in the analysis of individual aspects of the development of electromechanical systems. Some of these aspects are control design, electric circuit design and thermodynamics. A simulation tool applicable for the support of the design process of electromechanical systems should allow the simulation engineer to choose which physical effects he or she wants to consider in the simulation. Therefore, a simulation tool utilizing an open programming standard is more desirable than a proprietary tool. In this work the practical implementation of the proposed power electronic models is done with Modelica using the Dymola software platform.

Figure 4 shows the simulation of a battery powered air conditioning system (for heavy duty vehicles) written in Modelica. The compressor is driven by an induction machine with an integrated inverter and a control system. In order to generate an appropriate DC-link voltage for the inverter, the battery voltage gets boosted by a two-stage boost converter. Mechanical, electrical and thermal effects are considered in the model as shown in fig. 4. It can be used for choosing the core components of the air conditioning system and for calculating the overall system efficiency of the electrical components.

### 3. Object oriented modeling with Modelica

All models presented in this work have been implemented using Modelica. Modelica is an open programming standard and supports object orientation and multi-domain modeling. In Fritzson (2004) is described how to utilize Modelica to create models of any kind of physical object or device that can be described by algebraic equations and ordinary differential equations. Typical examples for such physical objects are electric components such as resistors, inductors, capacitors, semiconductors, etc. and mechanical components such as



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**Algorithm 1** Pseudo code of DC-to-DC converter models for calculating switching events in CICM.

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**Model:**

BuckConverter, BoostConverter, BuckBoostConverter

**Parameter:**

$L, C, R_S, R_L, R_D, V_D, f_s$

**Real variables:**

$V_{in}, V_{out}, i_S, i_L, i_D, i_{load}, t, d$

**Boolean variables:**

$s_{control}$

**Equations:**

**if** ( $s_{control} = true$ ),

consider equations corresponding to the equivalent circuit with the MOSFET in conducting state and the freewheeling diode in blocking state

**else**

consider equations corresponding to the equivalent circuit with the MOSFET in blocking state and the freewheeling diode in conducting state

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not necessary to calculate the switching events in the power electronic model as long as the relevant losses are considered.

## 5. Conventional models calculating switching events

It is widely accepted that power electronic circuits can be modeled with semiconductor models, that have a discrete blocking state and a discrete conducting state. This means that the respective semiconductor model has a very large off-resistance, a very low on-resistance and a knee voltage, if applicable (e.g. in diodes and IGBTs). Such semiconductor models are implemented with if-clauses and change their state at every switching event. Consequently, the entire circuit changes whenever one or more semiconductors change their states. Many simulation tools calculate the behavior of power electronic circuits using this approach or a similar one. This causes a considerable processing effort. The pseudo code in alg.1 can be used to implement models with discrete states of the circuits shown in fig. 1 – fig. 3. Usually switching losses are not considered in such models. However, the approach for the calculation of switching losses described in subsection 6.2 can also be applied to conventional models calculating switching events.

## 6. Averaged models with consideration of power balance

Converter models, built with the approach described in section 5, require a lot of processing time because the signal changes caused by switching events get calculated. If the calculation of switching events can be omitted the simulation time can be decreased significantly. Using the method of system averaging described by Cuk & Middlebrook (1978) it is possible to derive converter models without the calculation of switching events. In this work the general approach of system averaging is applied to the three DC-to-DC converter topologies shown in fig. 1 – fig. 3. For the calculation of the conduction losses in the averaged models presented in this work the ohmic contributions of the storage inductors  $R_L$ , the on-resistances of the MOSFETs  $R_S$ , and the on-resistances  $R_D$  and knee voltages  $V_D$  of the freewheeling diodes are considered. The presented models are valid for converter operation in continuous inductor current conduction mode (CICM).

### 6.1 Output voltage and conduction losses

In order to build an averaged model that considers power balance, a relation for the output voltage has to be found. For the buck converter the averaged output voltage can be found by

$$\bar{v}_{\text{out,buck}} = d(v_{\text{in}} - \bar{i}_L R_S + \bar{i}_L R_D + V_D) - (\bar{i}_L R_L + \bar{i}_L R_D + V_D) \quad (1)$$

whereas for the boost converter,

$$\bar{v}_{\text{out,boost}} = \frac{-v_{\text{in}} + \bar{i}_L R_L + \bar{i}_L R_D + V_D}{d - 1} - \frac{d(\bar{i}_L R_D + V_D - \bar{i}_L R_S)}{d - 1} \quad (2)$$

and for the buck-boost converter,

$$\bar{v}_{\text{out,buck-boost}} = \frac{\bar{i}_L R_D + V_D + \bar{i}_L R_L}{d - 1} - \frac{d(v_{\text{in}} - \bar{i}_L R_S + \bar{i}_L R_D + V_D)}{d - 1} \quad (3)$$

where  $\bar{v}_{\text{out}}$  is the average value of the output voltage,  $v_{\text{in}}$  stands for the input voltage,  $\bar{i}_L$  represents the average value of the inductor current, and  $d$  is the duty ratio of the converter. For calculating the conduction losses in the semiconductors and the storage inductors of the DC-to-DC converters a relation for the RMS values of the currents through the MOSFETs  $I_{S,\text{rms}}$ , the freewheeling diodes  $I_{D,\text{rms}}$ , and the storage inductors  $I_{L,\text{rms}}$  must be known. Assuming that the waveforms of the currents through the storage inductors are very close to a triangular shape (This is true for any of the presented DC-to-DC converter topologies designed with reasonable efficiency.) it is possible to derive

$$I_{S,\text{rms}} = \sqrt{d \left[ I_{L,\text{min}}^2 + I_{L,\text{min}} \Delta I_L + \frac{\Delta I_L^2}{3} \right]} \quad (4)$$

$$I_{D,\text{rms}} = \sqrt{(1 - d) \left[ I_{L,\text{max}}^2 - I_{L,\text{max}} \Delta I_L + \frac{\Delta I_L^2}{3} \right]} \quad (5)$$

and

$$I_{L,\text{rms}} = \sqrt{I_{S,\text{rms}}^2 + I_{D,\text{rms}}^2} \quad (6)$$

where

$$I_{L,\text{min}} = \bar{i}_L - \frac{\Delta I_L}{2} \quad (7)$$

and

$$I_{L,\text{max}} = \bar{i}_L + \frac{\Delta I_L}{2}. \quad (8)$$

(4) - (8) hold for all three described converter topologies. Though the relations for the averaged inductor current and the inductor current ripple depend on the converter topology. The averaged inductor current in the buck converter can be found by

$$\bar{i}_{L,\text{buck}} = \bar{i}_{\text{load}} \quad (9)$$

whereas in the boost converter

$$\bar{i}_{L,\text{boost}} = \frac{\bar{i}_{\text{load}}}{1-d} \quad (10)$$

and in the buck-boost converter

$$\bar{i}_{L,\text{buck-boost}} = \frac{\bar{i}_{\text{load}}}{1-d} \quad (11)$$

where  $\bar{i}_{\text{load}}$  is the average value of the load current.

The relations for the inductor current ripples in the three converters can be expressed by

$$\Delta I_{L,\text{buck}} = \frac{\bar{v}_{\text{out}} + \bar{i}_L R_D + V_D + \bar{i}_L R_L}{L} (1-d) T_{\text{switch}} \quad (12)$$

$$\Delta I_{L,\text{boost}} = \frac{v_{\text{in}} - \bar{i}_L R_L - \bar{i}_L R_S}{L} d T_{\text{switch}} \quad (13)$$

$$\Delta I_{L,\text{buck-boost}} = \frac{v_{\text{in}} - \bar{i}_L R_L - \bar{i}_L R_S}{L} d T_{\text{switch}} \quad (14)$$

where  $L$  is the storage inductance and  $T_{\text{switch}} = \frac{1}{f_{\text{switch}}}$  is the switching period.

The averaged input currents of the three converters can be calculated by

$$\bar{i}_{\text{in}} = \frac{P_{\text{out}} + P_{\text{con}} + P_{\text{switch}}}{v_{\text{in}}} \quad (15)$$

with

$$P_{\text{out}} = \bar{v}_{\text{out}} \cdot \bar{i}_{\text{load}} \quad (16)$$

and

$$P_{\text{con}} = R_S I_{S,\text{rms}}^2 + R_D I_{D,\text{rms}}^2 + V_D \bar{i}_D + R_L I_{L,\text{rms}}^2 \quad (17)$$

where  $P_{\text{con}}$  are the total conduction losses and  $P_{\text{switch}}$  are the total switching losses (calculated in subsection 6.2 by (21)). As the input voltage and the load current can be determined by simulating the respective converter model together with an appropriate voltage source model and load model, all quantities of the power balance are known.

## 6.2 Switching losses

The calculation of detailed voltage and current waveforms in a semiconductor during the transition from the blocking state to the conducting state and vice versa, caused by parasitic capacitances and inductances, takes a lot of processing time if differential equations are used to describe the waveforms during switching. Aubard et al. (2002) proposed a detailed dynamic model based on physical analysis of charge locations, that is too complex for multi-domain simulation. For models using piecewise, linear approximations such as presented by Eberle et al. (2008), many parameters must be known with good accuracy. Drofenik & Kolar (2005) presented a straight forward method to calculate switching losses

with polynomial approximation. This method is used (with small adaptations for MOSFET DC-to-DC converters) for the switching loss calculation in this work.

If the switching times are not used in a semiconductor model, it is possible to calculate the switching losses at various operation points using known switching loss characteristics (for the MOSFET and the freewheeling diode) together with the currents commutating between the MOSFET and the freewheeling diode  $i_{\text{switch on}}$  and  $i_{\text{switch off}}$ , the blocking voltage  $v_{\text{blocking}}$ , and the switching frequency  $f_{\text{switch}}$ . It is most practicable to generate the coefficients of the switching loss characteristics ( $a_n$ ,  $b_n$  and  $c_n$ ) from measurements of a ready made device at different commutating currents  $i_{\text{ref,switch on}}$  and  $i_{\text{ref,switch off}}$ . This is necessary because the switching losses are heavily dependent on the specific MOSFET-diode combination, the layout of the PCB and the gate resistance of the MOSFET. The model can be further improved if two switching loss characteristics (e.g.  $a_{n,50^\circ\text{C}}$  and  $a_{n,120^\circ\text{C}}$ ) can be generated from measurements at two different temperatures so that junction temperature dependence can be considered too. By measuring three operation points in a converter the quadratic switching loss characteristics

$$P_{S,\text{switch on}}(i_{S,\text{switch on}}) = a_1 i_{S,\text{switch on}} + a_2 i_{S,\text{switch on}}^2 \quad (18)$$

$$P_{S,\text{switch off}}(i_{S,\text{switch off}}) = b_1 i_{S,\text{switch off}} + b_2 i_{S,\text{switch off}}^2 \quad (19)$$

$$P_{D,\text{switch off}}(i_{D,\text{switch off}}) = c_1 i_{D,\text{switch off}} + c_2 i_{D,\text{switch off}}^2 \quad (20)$$

that specify the current dependence at a given switching frequency  $f_{\text{ref,switch}}$  and blocking voltage  $v_{\text{ref,blocking}}$  can be generated. The switching losses in the converter can be calculated from

$$P_{\text{switch}} = \frac{f_{\text{switch}}}{f_{\text{ref,switch}}} \frac{v_{\text{blocking}}}{v_{\text{ref,blocking}}} [P_{S,\text{switch on}}(i_{S,\text{switch on}}) + P_{S,\text{switch off}}(i_{S,\text{switch off}}) + P_{D,\text{switch off}}(i_{D,\text{switch off}})] \quad (21)$$

with the commutating currents found by

$$i_{S,\text{switch on}} = I_{L,\text{min}} \quad (22)$$

$$i_{S,\text{switch off}} = I_{L,\text{max}} \quad (23)$$

$$i_{D,\text{switch off}} = i_{S,\text{switch on}} \quad (24)$$

where the minimum and maximum value of the inductor current ripple  $I_{L,\text{min}}$  and  $I_{L,\text{max}}$  is given by (7) and (8) with the specific relations for each converter taken from (9)–(14). The different blocking voltages for the three converters are

$$v_{\text{blocking,buck}} = v_{\text{in,buck}} \quad (25)$$

$$v_{\text{blocking,boost}} = \bar{v}_{\text{out,boost}} \quad (26)$$

$$v_{\text{blocking,buck-boost}} = v_{\text{in,buck-boost}} + \bar{v}_{\text{out,buck-boost}} \quad (27)$$

where  $\bar{v}_{\text{out,boost}}$  and  $\bar{v}_{\text{out,buck-boost}}$  can be found by (2) and (3). From an analytical point of view, in (25)–(27) some very small voltage drops across the MOSFET and the freewheeling diode in on-state are neglected. However, in DC-to-DC converters with proper design and reasonable efficiency these voltage drops are very small compared to the input and output voltages.

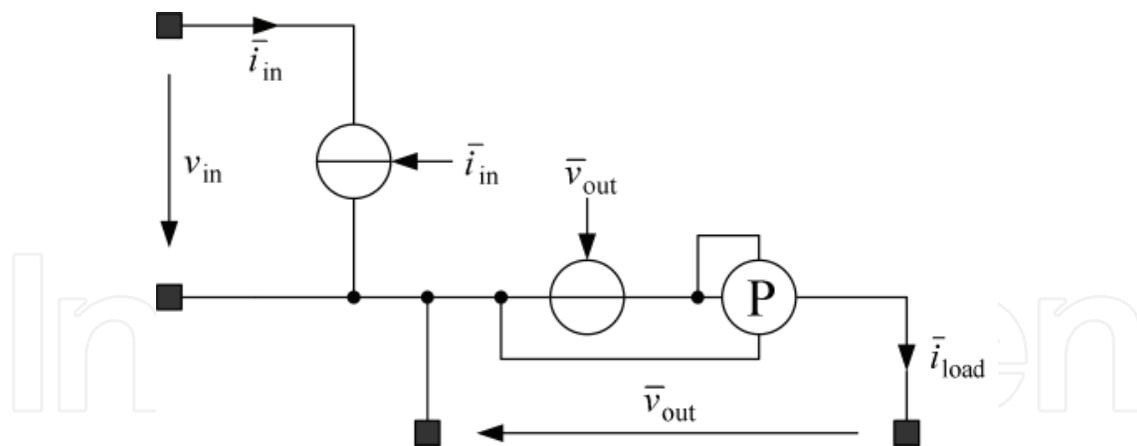


Fig. 5. Scheme of the averaged model of a DC-to-DC converter.

### 6.3 Temperature dependence

On-resistances and knee voltages in semiconductors as well as the ohmic contribution of the storage inductance change significantly when the junction temperature increases. Also switching losses are temperature dependent. Therefore it is useful to implement temperature dependence by linear or polynomial approximation (depending on the number of reference points). In this work linear temperature dependence is implemented for the on-resistance of the MOSFET and the freewheeling diode, the knee voltage of the freewheeling diode, the ohmic contribution of the storage inductance and for the reference points defining the current dependence of the switching losses. In a multi-domain simulation the presented loss models can be connected to thermal-network models of the converters. In such way it is possible to determine the thermal behavior of the converters by simulation.

### 6.4 Implementation of the averaged models

The basic scheme of the averaged model of a DC-to-DC converter, such as in fig. 1 – 3, is shown in fig. 5. The model of the energy source (e.g. a battery model) has to be connected to the two pins of the input side of the converter to define  $v_{in}$  and the model of the load (e.g. a DC-motor model) has to be connected to the output pins to close the circuit so that a current  $\bar{i}_{load}$  can flow. The output voltage is generated by the controlled voltage source where  $\bar{v}_{out}$  is calculated through the converter-specific relation among (1)–(3) and the input current is generated by the controlled current source where  $\bar{i}_{in}$  can be found by (15).

## 7. Simulation and comparison

For the verification of the proposed models the simulation results of a voltage controlled buck converter modeled with the calculation of switching events (as in section 5), indicated as model 1, and modeled as an averaged system (as in section 6), indicated as model 2 are compared. The buck converter in the two models operates with a switching frequency of 50 kHz and a maximum output power of 120 W. The storage inductance is 300  $\mu$ H and the buffer capacity is 200  $\mu$ F.

In fig. 6 and 7 screen shots of model 1 and model 2 are shown. It can be seen in fig. 7 that the capacitor at the input of the converter is not modeled. This is because in model 2 the mean value of the input current gets calculated instead of the actual pulsed input current calculated in model 1.

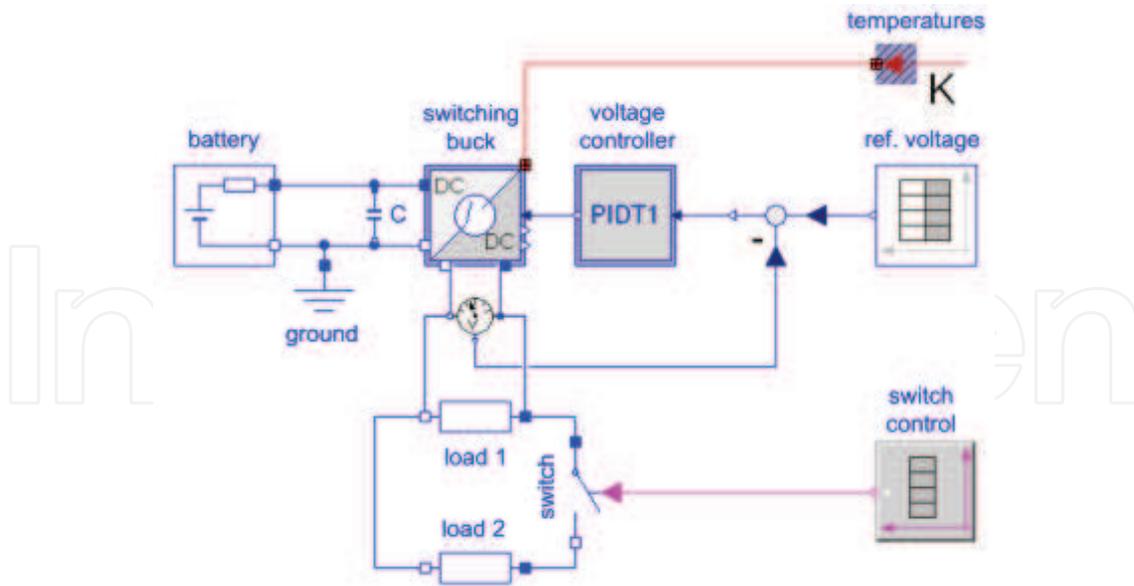


Fig. 6. Model of a voltage controlled buck converter with calculation of switching events (model 1).

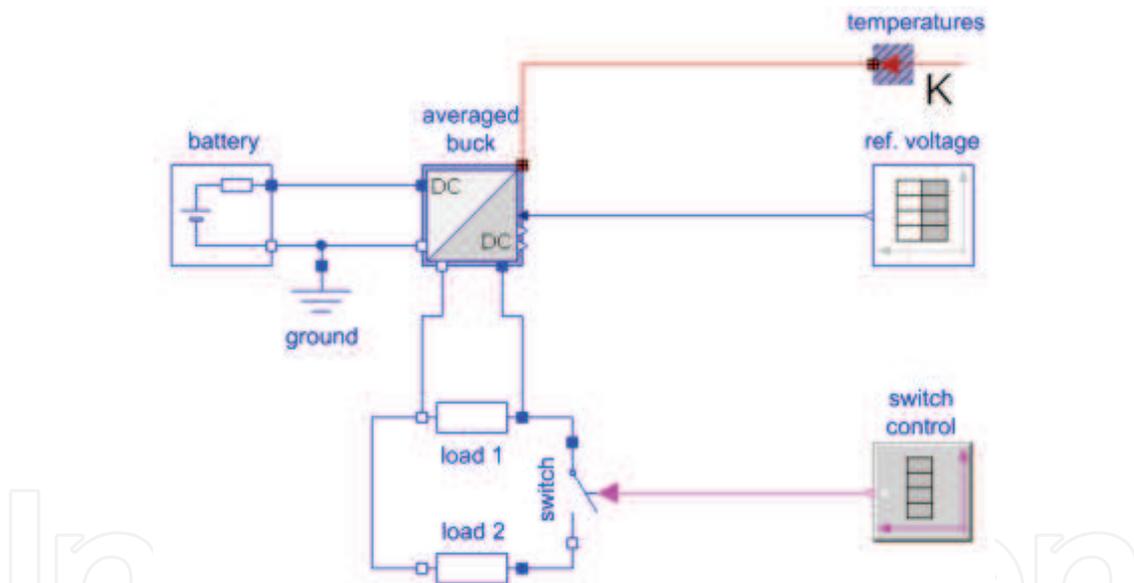


Fig. 7. Averaged model of a voltage controlled buck converter (model 2).

Furthermore, there is no linear voltage controller modeled in fig. 7 because in model 2 only steady states of the converter are calculated. Therefore, an inverse model of (1) for the calculation of the duty cycle is sufficient. This inverse model is implemented directly in the buck converter model. Instead of the differential equations of a controller only an algebraic equation has to be solved.

Both models were processed on a conventional personal computer with a 2.4 GHz dual core CPU (However, only a single core was used for processing.) and 3.0 GB RAM. The simulation of model 1 takes 2.89 s CPU time whereas the simulation of model 2 takes only 15 ms. This shows that the processing of model 2 is significantly faster than the processing of model 1.

Some simulation results of model 1 and model 2 are shown in fig. 8–10. A reference voltage step of 20 % and a load increase of 100 % are simulated with both models. The voltage step occurs between time 0.01 s and time 0.03 s, and the load increases after the switch between the two equal load resistances closes at time 0.05 s.

In fig. 8 the influence of the converter input current on the battery voltage can be observed. The output voltage waveforms match in steady state. (The ripple of the output voltage of model 1 is too small to be recognized.) During changes of the operation point the influence of the voltage controller and the transient behavior of the converter can be seen in the output voltage of model 1. These transient effects appear also in the current and power signals of model 1 in fig. 9 and 10. In the input current and inductor current waveforms of model 1 the switching of the converter becomes obvious. Please note, that in fig. 10 the signals of model 1 are mean values averaged over a switching period  $T_{\text{switch}}$ .

The mean value of any steady state voltage or current signal of model 2 matches the corresponding signal in model 1. Consequently, also the steady state results of the input power, and the total converter losses match. Therefore, model 1 and model 2 have an equivalent steady state behavior.

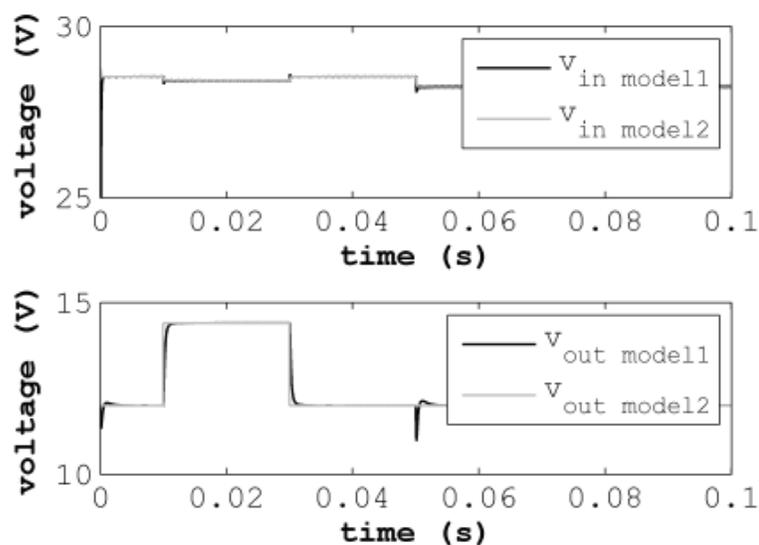


Fig. 8. Voltage waveforms in model 1 and model 2. Input voltages (top); Output voltages (bottom).

## 8. Conclusion

A computation time efficient method using system averaging and polynomial approximation for modeling three DC-to-DC converters with consideration of temperature dependent conduction and switching losses is described and implemented. The models are developed for multi-domain simulations of electromechanical systems. These simulations are used to investigate system efficiency taking electrical, mechanical and thermal effects into account while assuring power balance. The simulation results of two voltage controlled buck converter models are presented and compared. Voltages and currents calculated with the

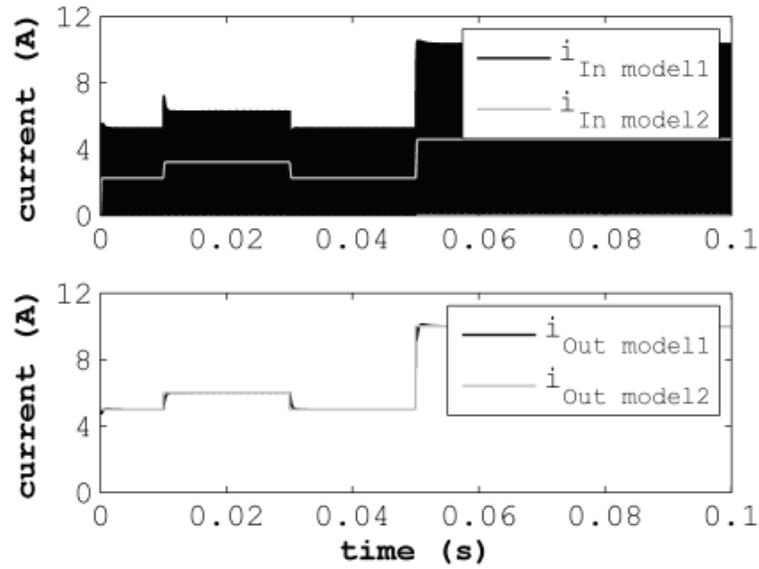


Fig. 9. Current waveforms in model 1 and model 2. Input currents (top); Output currents (bottom).

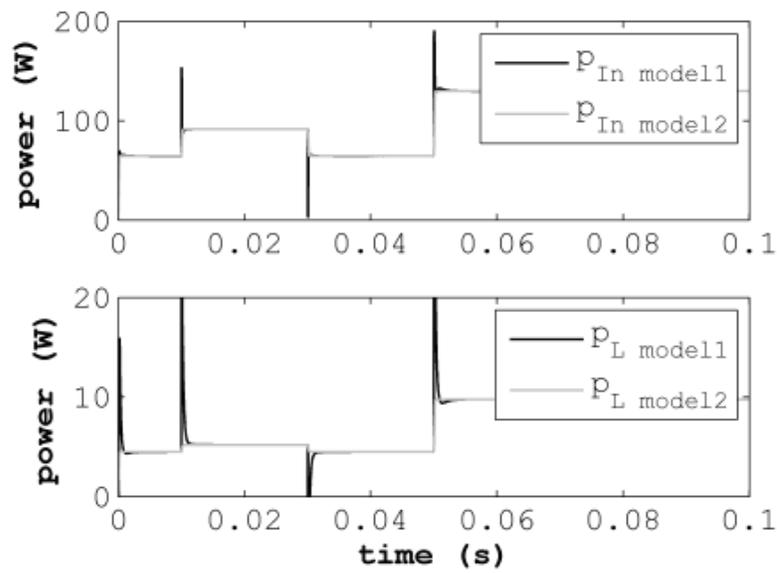


Fig. 10. Power in model 1 and model 2. The signals of model 1 are mean values. Input power (top); Total converter losses (bottom).

averaged model (model2) match the respective mean values of the conventional model (model 1) in steady state.

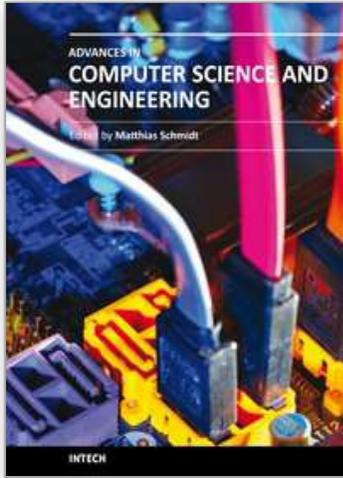
A comparison of the CPU times of the two presented simulations shows that the proposed averaged models can be processed significantly faster than conventional models describing semiconductors with discrete states.

## 9. Acknowledgment

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