Modeling Controlled Switches and Diodes for Electro-Thermal Simulation

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Abstract – **Designers of advanced power converters may choose from a variety of switching device models for simulation. Some situations call for simple idealized models, while others require physics-based models. When evaluating thermal system performance, a behavioral model that includes both conduction and switching losses is desired. A set of models has been developed to include both unidirectional devices, such as IGBTs, BJTs, and diodes, and bidirectional devices, such as MOSFETs. Logic and timing elements are used to insert voltage and current sources into the circuit at appropriate times. All losses affect circuit operation, so simulation can accurately predict losses when the load affects the switching pattern. The model was constructed in Dymola and included thermal ports to be attached to a model of the thermal system. Temperature dependency of device parameters can be included with minor modifications. Experimental verification is shown.**

I. Introduction

Models used in power electronics simulation are typically either extremely complex dynamic models or simple conduction loss models. Electro-thermal simulation calls for a model of moderate complexity. Here a new model is derived to include switching losses in addition to conduction losses in an IGBT (insulated gate bipolar transistor) or MOSFET based switching pole.

Most published switch models are fundamentally physics-based. A comprehensive review of IGBT models is available [1], although new models are still being developed. A physics model is extended in [2] to include thermal effects. There are also a few published behavioral models. A model is built in Simulink with a parallel capacitor to model switching behavior in [3]. The method in [4] is similar to [3] but uses a nonlinear capacitor to model switching behavior more accurately. A more complex system developed in [5] attempts to model each voltage and current transition in a switching event with piecewise linear *dv/dt* and *di/dt*. Switching losses in MOSFETs have been discussed in [6]. Each of these relies on detailed information regarding switching transients, perhaps measured by an oscilloscope. Unfortunately, such measurements are useful for voltage transient studies, but are notoriously unreliable as power measurements due to limited dynamic range.

The model developed focuses on power dissipation, a problem more central to the design of a typical power converter than voltage transients. Averaging is used in [7] to create a power dissipation model. The present work instead inserts power dissipating elements at each switching edge, such that any method of pulse generation may be used. The method is therefore highly appropriate for non-periodic switching schemes, such as hysteresis current control, delta modulation current control, non-standard space vector modulation, or direct torque control. The model can also be used for certain switched reluctance motor drives in which the modulation scheme changes for different portions of the cycle.

II. IGBT/Diode Model Development

Previous work defined IGBT/diode models [8]. The development that follows is performed for a "buck-derived" switch, shown in Fig. 1. The switch pole has two ports, one defined primarily by voltage and one defined by current (positive going out). If the controlled switch and diode are swapped, the following discussion also applies, with appropriate logic and polarity changes.

Consider the idealized waveforms of Fig. 2 showing a typical IGBT and its soft free-wheeling diode. There are two switching events, with wide voltage swings and current commutation, separated by two conduction periods. Including all of these dynamics in a simulation would result in excessively long simulation times. The time step would need to be small enough to capture all of the *dv/dt* and *di/dt* slopes. Also, simulation methods for controlling the slopes are difficult at best, usually relying on a capacitance methods [3],[5].

If we assume that switching energy varies linearly with current, the energy dissipated at each switching instant is:

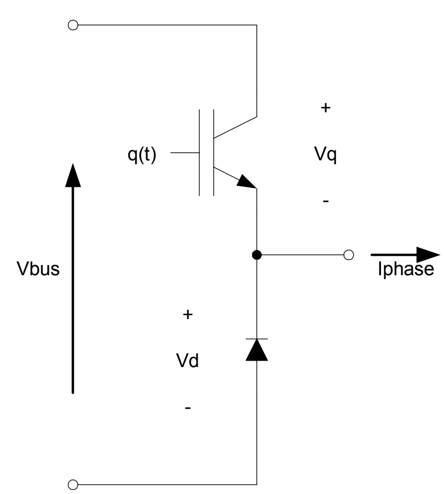

Fig. 1. Buck-derived switch pole.

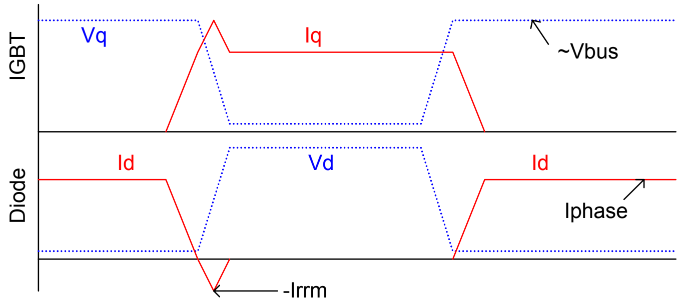

Fig. 2. Idealized switching waveforms.

$$E = kV_{bus} I_{phase} t_{sw} \tag{1}$$

where V_{bus} and I_{phase} are defined in Fig. 1, t_{sw} is a characteristic switching time, and k is a proportionality constant. Arbitrary power waveforms, including triangle and half-sinusoid power pulses, are often modeled with a rectangle whose height is 70% of the peak power, and whose base is an appropriate width to give equal energy. Given E, V_{bus}, and I_{phase}, and setting $k=0.7$, t_{sw} can be calculated. Generally, E can represent E_{on} (turn-on energy) or E_{off} (turn-off energy) or E_{sw} (total switching energy). The model developed separates E_{on} from E_{off}, if the parameters can be obtained by some means.

The definition of a model follows directly from (1). A voltage-controlled voltage source with gain k and control input V_{bus} can be inserted in series with I_{phase} for time t_{sw}. This results in the voltage waveforms shown in Fig. 3. The phase voltage decreases in steps rather than linearly, simplifying the simulation. The simulator need only include the corner points, rather than a large number of points that define switch transitions.

Standard diodes (including ultrafast, ultrasoft, etc.) also dissipate energy due to reverse recovery:

$$E_{rec} = \frac{kI_{rrm} V_{bus} t_{rr}}{2} \tag{2}$$

The factor of ½ results from the transition of the terminal voltage during the reverse-recovery period. Again $k=0.7$, using the same reasoning as above. An equal amount of energy also adds to turn-on losses in the IGBT, so the complete form for E_{on} is:

$$E_{on} = kV_{bus} I_{phase} t_{sw} + \frac{kI_{rrm} V_{bus} t_{rr}}{2} \tag{3}$$

The reverse-recovery phenomenon can be modeled as a current source of magnitude kI_{rrm} connected across the bus terminals, which is only turned on for time t_{rr}. The power consumed in this current source is then apportioned equally to the IGBT and diode. As above, the preferred method is to measure energy and deduce t_{rr}.

The complete model includes both conduction loss and the new switching loss terms. The voltage-controlled voltage source and current source are turned on using one-shot elements and logic. Complete Dymola/Modelica models are available [9].

III. MOSFET Model Development

The IGBT/diode derivation in section II relies on a fundamental assumption: device current is unidirectional. For most controlled switches and for diodes, unidirectional current is assured. However, for MOSFETs, channel conduction is fundamentally bidirectional, and the chip includes an integral body diode. Clearly a change in logic is necessary.

A MOSFET switching pole is shown in Fig. 4. Symmetrical construction with synchronous rectification is assumed. By allowing the parameters to vary, and allowing one switch to be always off, a lower performance system that does not use synchronous rectification can also be modeled. An extra Schottky diode is added in parallel in many practical converters in order to prevent the body diode from conducting. Schottky diode switching characteristics are far superior to MOSFET body diodes, with effectively no stored charge and no reverse recovery. The logic discussed below covers this application; the Schottky diode model only includes conduction loss and is external to the MOSFET switching pole.

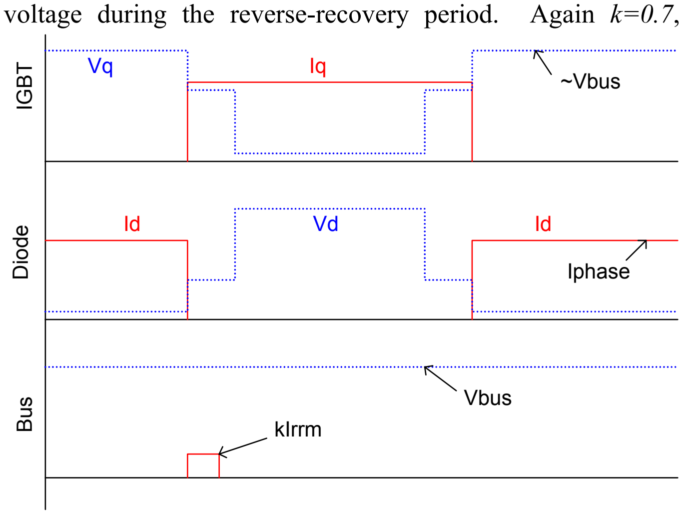

Fig. 3. Modeled waveforms.

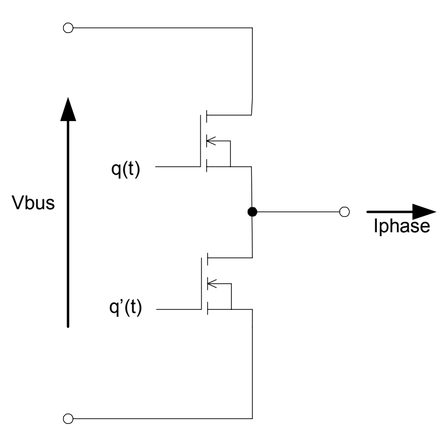

Fig. 4. MOSFET switching pole

The internal model of the MOSFET switching pole is shown in Fig. 5. The channel is modeled as a switch with on-state resistance; the diode is modeled with an on-state voltage and on-state resistance. Either MOSFET may incur commutation losses, so two switched voltage sources are included. Either diode may experience reverse recovery, so two switched current sources are included. Partial Modelica code is also given in the appendix; a full listing is available [9].

The most important variables define the switch states just prior to switch action: diode currents, channel currents, and device voltages. Dymola offers a function, $pre(x)$, whose value is the left-limit of x at a given instant:

$$pre\left(x(t)\right) = \lim_{t' \to t^-} x(t') \tag{4}$$

Using $pre(x)$ in conjunction with edge detection via a "when" statement creates a sample-and-hold of the relevant current or voltage, which is then used to determine which sources, if any, must be activated, as shown in Table I.

A more realistic model for diode reverse recovery would set the value of I_{rr1} and I_{rr2} to be a function of the previous conducted current. This proposed function is difficult to measure in a power converter. In practice, the reverse-recovery current is usually estimated from datasheet values. In the interest of simplicity, a step function is used in the present model. A reasonable threshold is the corner of the piecewise-linear conduction model, I_x, which essentially defines the point of full diode conduction. The current in the MOSFET body diode should be below the threshold if a reasonably chosen Schottky diode is added in parallel, thus eliminating the body diode's reverse-recovery from the simulation and from circuit operation.

Simulated waveforms are shown in Fig. 6. A conduction-loss-only model is compared to a full model. Because of the high overall efficiency, differences in the current are difficult to discern, but differences in power dissipation are clearly visible on the temperature plots.

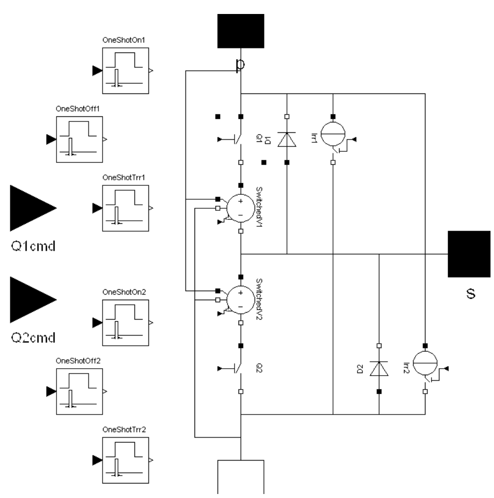

Fig. 5. Dymola MOSFET model, schematic view.

TABLE I. RELEVANT STATES FOR MOSFET MODEL

Switch Edge	Relevant Pre Variable	Condition	Source to Activate
Q1 on	V(Q1)	Pre(V(Q1)) > 0	SwitchedV1
Q1 off	I(Q1)	Pre(I(Q1)) > 0	SwitchedV1
Q2 on	V(Q2)	Pre(V(Q2)) > 0	SwitchedV2
Q2 off	I(Q2)	Pre(I(Q2)) > 0	SwitchedV2
Q1 on	I(D2)	Pre(I(D2)) > Ix	Irr2
Q2 on	I(D1)	Pre(I(D1)) > Ix	Irr1

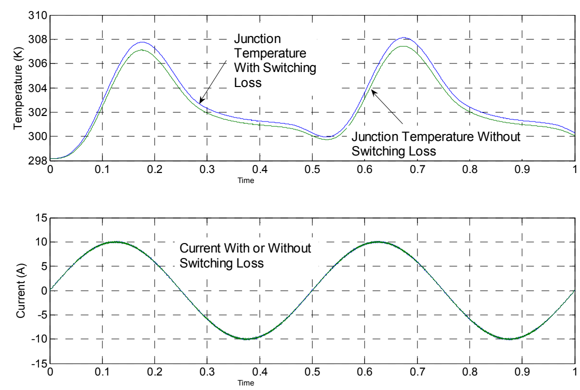

Fig. 6. Simulated MOSFET current and temperature.

IV. CALORIMETRY FOR MODEL PARAMETERIZATION

The models developed above rely on information about power dissipation. Many methods have been used to measure power loss in a switching pole. Conduction loss measurements are straightforward; a high-power curve tracer such as a TEK371 can be used. Switching loss is much more difficult to measure. Oscillograms of current and voltage can be obtained and integrated, but not without disturbing the circuit. Finite resolution and bandwidth further reduce the accuracy of this method.

The standard method for measuring losses in highly efficient converters is calorimetry [10]. The basic concept is to measure the temperature difference across a known thermal resistance. Thermal resistance is difficult to compute, so it is normally measured using a calibration resistor whose power dissipation can be accurately measured electrically.

A new variation on calorimetry was developed to speed the process, shown in Fig. 7. Only the outer insulated box was used in [10]. This gives high accuracy due to high thermal resistance, but also requires an inordinate amount of time. Good insulators have both low thermal conductivity and high specific heat, for long thermal time constants. The box must reach thermal equilibrium for a valid test point, which may take more than a day. The method used in the present work instead uses a double box, where the inner box is corrugated cardboard and the outer box is heavily insulated. The outer box provides a stable environment with immunity from room

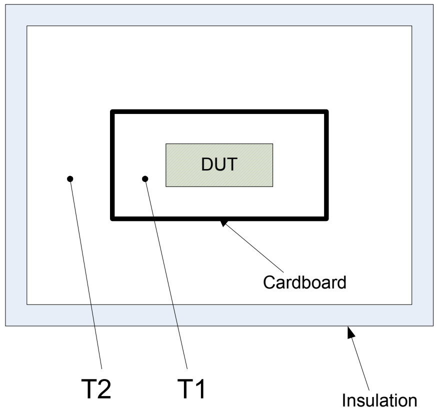

Fig. 7. Modified calorimeter.

temperature transients, while the inner box provides the thermal resistance that is measured. The relevant time constant is approximately 45 min. due to the low thermal mass of the cardboard box.

Several tests were run, some with the IGBT/diode switching converter operating, some with a calibration resistor. Prior to assembly, the switching devices were characterized on a TEK371 to find fixed-voltage-plus-fixed-resistance models:

$$\begin{aligned} V_{IGBT}(I) &= V_{on,Q} + R_{on,Q} I \\ V_{FWD}(I) &= V_{on,D} + R_{on,D} I \end{aligned} \tag{5}$$

Test points at fixed frequency were used to find model parameters from:

$$\begin{aligned} P_{conduction} &= \left(V_{on,Q} D + V_{on,D}(1-D)\right) I + \cdots \\ &\quad \left(R_{on,Q} D + R_{on,D}(1-D)\right) I^2 \\ P_{switching} &= cf + dfI \\ c &= kV_{bus} I_{rrm} t_{rr} \\ d &= 2kV_{bus} I_{phase} t_{sw} \\ P_{measured} &= P_{conduction} + P_{switching} \end{aligned} \tag{6}$$

Experimental data is used to find parameters c and d. Switching time t_{sw} can be found directly from the definition of d. For reverse recovery, only the total ampere-seconds can be determined from power measurements. In the present work, t_{rr} was estimated from manufacturer's data and I_{rrm} was derived from c and t_{rr}. Experiments with MOSFETs are in progress.

V. Experimental Results

A calorimetry calibration curve is shown in Fig. 8. Note the high degree of linearity and tight correlation. This curve is used to calculate power dissipation for the switching converter.

Model extraction test points were based on hysteresis current control (see Fig. 9). With a fixed load, hysteresis current control will result in fixed frequency and duty cycle. Equation (6) was used to find model parameters, shown in Table II. These parameters were used in Dymola.

Next, delta current modulation was used. In delta modulation, a fixed clock samples a comparison between actual current and a reference. The resulting switching waveform typically exhibits subharmonics and other aperiodic behavior. As with hysteresis control, waveform parameters depend on both the source and the load, but in a more complex manner. The power dissipation associated with this modulation scheme is difficult or impossible to estimate analytically, so a Dymola simulation was used for comparison to experimental data. Results are summarized in Table III.

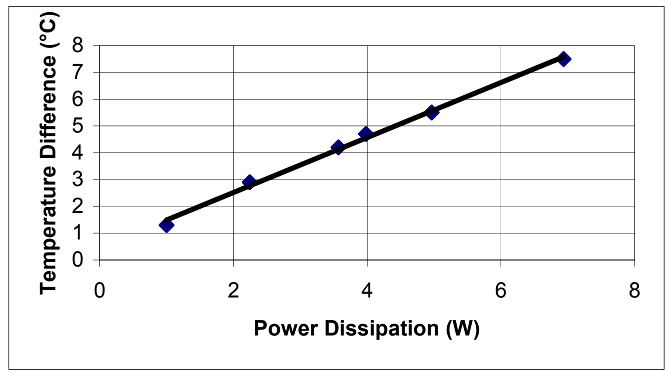

Fig. 8. Calibration curve for calorimeter.

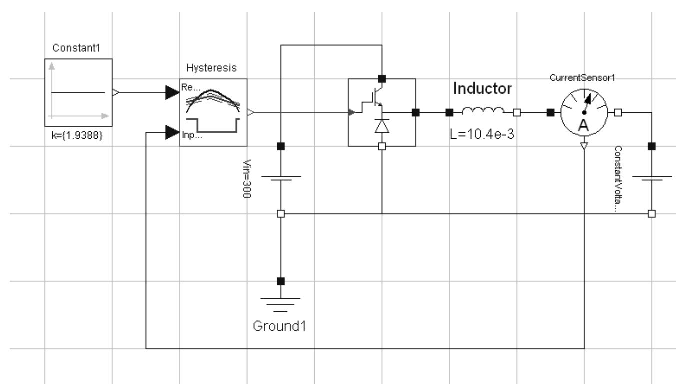

Fig. 9. Hysteresis current controlled converter.

TABLE II. MODEL FROM EXPERIMENTS

$V_{on,Q}$	0.940 V	$V_{on,D}$	0.500 V
$R_{on,Q}$	0.105 W	$R_{on,D}$	0.050 W
t_{on}	236 ns	t_{rr}	38.5 ns
t_{off}	236 ns	I_{rrm}	2.018 A

TABLE III. EXPERIMENTAL VS. SIMULATED DISSIPATION
14.9 KHZ DELTA MODULATION, 300 V INPUT, 70 V OUTPUT

Phase Current	Measured Dissipation	Simulated Dissipation	Error
1.840 A	2.08 W	2.09 W	0.48%
3.699 A	4.91 W	5.05 W	2.85%
4.975 A	6.85 W	6.95 W	1.46%

VI. Conclusions

A new model of a switching converter has been developed, which can be used to estimate power dissipation due to both conduction and switching loss. The model is suitable for all modulation schemes, including those with no fixed frequency. Model extraction based on a modified calorimetry scheme has been demonstrated, using hysteresis current control and delta current modulation.

References

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Appendix: Partial Modelica Listing of MOSFET Model

```
model Switching "MOSFET Bridge with Switching
Losses"
  parameters …;
  components …;
equation
  Vx = VonD;
  Ix = VonD/Rdson;
  Q1bar = not (Q1cmd.signal[1]);
  Q2bar = not (Q2cmd.signal[1]);
  PQ1 = (Q1.i*Q1.v) + (D1.i*D1.v) +
(Q1.i*SwitchedV1.v) + 0.5*(Irr1.i + Irr2.i)
    *(p.v - n.v);
  PQ2 = (Q2.i*Q2.v) + (D2.i*D2.v) +
(Q2.i*SwitchedV2.v) + 0.5*(Irr1.i + Irr2.i)
    *(p.v - n.v);
  OneShotOn1.inPort.signal[1] = Q1cmd.signal[1];
  OneShotOff1.inPort.signal[1] = not
(Q1cmd.signal[1]);
  OneShotTrr1.inPort.signal[1] = Q2cmd.signal[1];
  OneShotOn2.inPort.signal[1] = Q2cmd.signal[1];
  OneShotOff2.inPort.signal[1] = not
(Q2cmd.signal[1]);
  OneShotTrr2.inPort.signal[1] = Q1cmd.signal[1];
  when edge(Q1cmd.signal[1]) then
    preID2 = pre(D2.i);
    preVQ1 = pre(Q1.v);
  end when;
  when edge(Q2cmd.signal[1]) then
    preID1 = pre(D1.i);
    preVQ2 = pre(Q2.v);
  end when;
  when edge(Q1bar) then
    preIQ1 = pre(Q1.i);
  end when;
  when edge(Q2bar) then
    preIQ2 = pre(Q2.i);
  end when;
  Irr1.TurnOn.signal[1] =
OneShotTrr1.outPort.signal[1] and (preID1 > Ix);
  Irr2.TurnOn.signal[1] =
OneShotTrr2.outPort.signal[1] and (preID2 > Ix);
  SwitchedV1.TurnOn.signal[1] =
(OneShotOff1.outPort.signal[1] and (preIQ1 > 0))
     or (OneShotOn1.outPort.signal[1] and (preVQ1 >
0));
  SwitchedV2.TurnOn.signal[1] =
(OneShotOff2.outPort.signal[1] and (preIQ2 > 0))
     or (OneShotOn2.outPort.signal[1] and (preVQ2 >
0));
  Q1.s.signal[1] = Q1cmd.signal[1] or
OneShotOff1.outPort.signal[1];
  Q2.s.signal[1] = Q2cmd.signal[1] or
OneShotOff2.outPort.signal[1];
  connections…;
end Switching
```