Power Electronics Circuits

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ECE 482 Lecture 4
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Simulation Modeling
Circuit Simulation

- LTSpice
  - Other tools accepted, but not supported
- Choose model type (switching, averaged, dynamic)
- Supplement analytical work rather than repeating it
- Show results which clearly demonstrate what matches and what does not with respect to experiments (i.e. ringing, slopes, etc.)

LTSpice Modeling Examples

- Example files added to course materials page
Custom Transistor Model

```verbatim
.model myD D(n=.001)  
.model mySw SW(Ron=10m Roff=1G Vt=0 Von=1 Voff = .5 )
```

VDMOS Model

<table>
<thead>
<tr>
<th>Name</th>
<th>Description</th>
<th>Units</th>
<th>Default</th>
<th>Example</th>
</tr>
</thead>
<tbody>
<tr>
<td>Vto</td>
<td>Threshold voltage</td>
<td>V</td>
<td>0</td>
<td>1.0</td>
</tr>
<tr>
<td>Rn</td>
<td>Transconductance parameter</td>
<td>A/V 2</td>
<td>1.0</td>
<td>.5</td>
</tr>
<tr>
<td>Phi</td>
<td>Surface inversion potential</td>
<td>V</td>
<td>0.6</td>
<td>0.65</td>
</tr>
<tr>
<td>LndA</td>
<td>Channel-length modulation</td>
<td>I/V</td>
<td>0.002</td>
<td></td>
</tr>
<tr>
<td>gtride</td>
<td>Conductance multiply in triode regions independent for triode and saturation regions</td>
<td>-</td>
<td>1.0</td>
<td>2.0</td>
</tr>
<tr>
<td>subtitle</td>
<td>Current (gms volt Vds) to switch from square law to approximated subthreshold conduction</td>
<td>A/V</td>
<td>0.0</td>
<td>1.0</td>
</tr>
<tr>
<td>BV</td>
<td>Vds breakdown voltage</td>
<td>V</td>
<td>Inf</td>
<td>40</td>
</tr>
<tr>
<td>IMV</td>
<td>Current at Vds (mA)</td>
<td>A</td>
<td>100mA</td>
<td>1uA</td>
</tr>
<tr>
<td>IMV</td>
<td>Vds breakdown emission coefficient</td>
<td>-</td>
<td>1.0</td>
<td></td>
</tr>
<tr>
<td>Rds</td>
<td>Drain ohmic resistance</td>
<td>$\Omega$</td>
<td>0.0</td>
<td>1.0</td>
</tr>
<tr>
<td>Rs</td>
<td>Source ohmic resistance</td>
<td>$\Omega$</td>
<td>0.0</td>
<td>1.0</td>
</tr>
<tr>
<td>Rg</td>
<td>Gate ohmic resistance</td>
<td>$\Omega$</td>
<td>0.0</td>
<td>2.0</td>
</tr>
<tr>
<td>Rds</td>
<td>Drain-source shunt resistance</td>
<td>$\Omega$</td>
<td>Inf</td>
<td>10M$\Omega$</td>
</tr>
<tr>
<td>Rds</td>
<td>Body ohmic resistance</td>
<td>$\Omega$</td>
<td>0.0</td>
<td>.5</td>
</tr>
<tr>
<td>Cgs</td>
<td>Source-body diode</td>
<td>$F$</td>
<td>0.0</td>
<td>1n</td>
</tr>
</tbody>
</table>

http://ltwiki.org/LTspiceHelp/LTspiceHelp/M_MOSFET.htm
Manufacturer Device Model

- Text-only netlist model of device including additional parasitics and temperature effects
- May slow or stop simulation if timestep and accuracy are not adjusted appropriately

Full Switching Simulation
Full Switching Simulation

Experiment Procedure

- Problem Description
- Impedance and Resistance
- Experiment 3 Components
- Contents of the Experiment 3 Parts Kit
- Contents of the Magentic Library
- Breadboard Board Assembly and Layout

Reference Materials

- Power Converter Layout
- References of Analog Circuit Design (ZT App. Note)
- PCB Layout of Components: Obsevcd Wire Etch

LTSpice Example Files

- Example of power semiconductor modeling using custom or manufacturer models
- Example: Switching and Control LTSpice Models
- Additional example on integrated circuit modeling in LTSpice Equations
- Example: A single-transistor boost converter schematic diagram (Fig. A-1.1)
Switching Model Simulation Results

- Simulation Time ≈ 15 minutes

Full Switching Model

- Gives valuable insight into circuit operation
  - Understand expected waveforms
  - Identify discrepancies between predicted and experimental operation
- Slow to simulate; significant high frequency content
- Cannot perform AC analysis
Averaged Switch Modeling: Motivation

• A large-signal, nonlinear model of converter is difficult for hand analysis, but well suited to simulation across a wide range of operating points
• Want an averaged model to speed up simulation speed
• Also allows linearization (AC analysis) for control design

Averaged, Nonlinear, Large-Signal Equations

ECE 481 Review:

\[ L \frac{dv(t)}{dt} + v_L(t) - v(t) \]

\[ C \frac{di(t)}{dt} + i(t) \]

\[ \langle v(t) \rangle = \langle v_g(t) \rangle - \langle v_L(t) \rangle - \langle v(t) \rangle \]

\[ \langle i(t) \rangle = \frac{1}{R} \langle v(t) \rangle + \langle i(t) \rangle \]

DC: \[ \phi = V_g - V \]

AC: \[ C \frac{dv(t)}{dt} = i(t) - i(t) - \frac{v(t)}{R} \]
Nonlinear, Averaged Circuit

\[
L \frac{d\langle i_L \rangle}{dt} = \langle v_{bus} \rangle - (1-d)\langle v_{bus} \rangle
\]

\[
C \frac{d\langle v_{bus} \rangle}{dt} = (1-d)\langle i_L \rangle - \langle i_{bus} \rangle
\]

Implementation in LTSpice

\[
\langle v_1(t) \rangle_{T_2} = d\langle v_1(t) \rangle_{T_3} \\
\langle v_2(t) \rangle_{T_2} = d\langle v_2(t) \rangle_{T_3} \\
\langle i_1(t) \rangle_{T_2} = d\langle i_1(t) \rangle_{T_3} \\
\langle i_2(t) \rangle_{T_2} = d\langle i_2(t) \rangle_{T_3}
\]
Averaged Switch Model

Averaged Model With Losses

What known error(s) will be present in loss predictions with this model?
Experiment 4

Experiment 4: Closed-Loop Boost
Current Control

Current Programmed Control (CPM)

The peak transistor current replaces the duty cycle as the converter control input.
Current Programmed Control

• Covered in Ch. 12 of *Fundamentals of Power Electronics*
• Advantages of current programmed control:
  − Simpler dynamics — inductor pole is moved to high frequency
  − Simple robust output voltage control, with large phase margin, can be obtained without use of compensator lead networks
  − It is always necessary to sense the transistor current, to protect against overcurrent failures. We may as well use the information during normal operation, to obtain better control
  − Transistor failures due to excessive current can be prevented simply by limiting $i_c(t)$
  − Transformer saturation problems in bridge or push-pull converters can be mitigated
• A disadvantage: susceptibility to noise
The First-Order Approximation

\[ \langle i_L(t) \rangle_{T_s} = i_c(t) \]

- Neglects switching ripple
- Yields physical insight and simple first-order model
- Accurate when converter operates well into CCM (so that switching ripple is small)
- Accurate when artificial ramp (discussed later) small
- Resulting small-signal relation:
  \[ \hat{i}_L(s) \approx \hat{i}_c(s) \]

Averaged Modeling

from before:

1. \( L \frac{di_L}{dt} = \dot{v}_g - \dot{v}_c + \sqrt{2} \)
2. \( C \frac{d\theta_c}{dt} = -I_L \hat{d} + \dot{v}_c - \frac{\theta_c}{L} \)

Apply \( \dot{f}_L = \dot{f}_c \) into 1

Solve 1 for \( \dot{d} \) and plug into 2

\[ \dot{d} = \frac{1}{V} \left( sL\dot{f}_c - \dot{v}_g + \dot{v}_c \right) \]

Duty cycle \( \dot{d} \) is determined by: \( \dot{f}_c, \dot{v}_c, \dot{v}_g \)

3. \( C \frac{d\hat{v}_c}{dt} = \frac{V}{V} \left( sL\dot{f}_c - \dot{v}_g + \dot{v}_c \right) + \dot{v}_c - \frac{\dot{\theta}_c}{L} \)

Apply 4

\[ I_L = \frac{V - \dot{v}_c}{R} \]

\[ \dot{v}_c = \frac{V}{R} \]
Large-Signal Nonlinear Model

\[ C \frac{d^2 \phi}{dt^2} = -\frac{1}{R_b'} \left( sL \dot{i}_c - \dot{v}_b + D \dot{v} \right) + B \dot{i}_c + \frac{D}{R} \]

\[ C \frac{d\phi}{dt} = \dot{i}_c \left( B' - \frac{g_{m}}{R_b'} \right) + \frac{g_{m}}{R_b'} - \frac{\phi'}{R} \]

**AC equivalent circuit model:**

**Large-signal nonlinear model**

**Averaged switch network**

\[ i_2 = \frac{v_1 + i_2}{v} \]
Implementation in LTSpice

Perturb and Linearize
Boost CCM CPM Small-Signal Model

\[ G_{we} = \frac{\omega}{\xi_v} \bigg|_{\delta = 0} = D \left(1 - \frac{R}{D^2 R}\right) \frac{R}{1 + sC} \]

\[ = D \left(1 - \frac{R}{D^2 R}\right) \frac{R/2}{1 + sC/2} \]

Same RHP zero from Boost control

Single Pole!

Still limit BW to \( \omega \frac{R}{C} \) (\( \frac{R}{C} \) aggressive)

CPM Transfer Functions

Fig. C.23. Comparison of CPM control with duty-cycle control, for the control-to-output frequency response of the buck-boost converter example.
The current programmed controller is inherently unstable for $D > 0.5$, regardless of the converter topology.

Controller can be stabilized by addition of an artificial ramp.

CPM Oscillations for D>0.5
Inductor Current Waveform in CCM

Inductor current slopes $m_1$ and $-m_2$

- back converter
  $$m_1 = \frac{v_1 - v_2}{L}$$
  $$m_2 = \frac{v_2 - v_1}{L}$$
- boost converter
  $$m_1 = \frac{v_1}{L}$$
  $$m_2 = \frac{v_2}{L}$$
- buck-boost converter
  $$m_1 = \frac{v_1}{L}$$
  $$m_2 = \frac{v_2}{L}$$

Volt-Second Balancing

First interval:
$$i_2(dT_s) = i_2 - i_2(0) + m_1 dT_s$$

Solve for $d$:
$$d = \frac{i_2 - i_2(0)}{m_1 T_s}$$

Second interval:
$$i_2(T_s) = i_2(dT_s) - m_1 dT_s$$
$$= i_2(0) + m_1 dT_s - m_1 dT_s$$

In steady state:
$$0 = M_2 DT_s - M_1 DT_s$$
$$\frac{M_2}{M_1} = \frac{D}{D}$$
Introducing a Perturbation

Change in Inductor Current Over $T_s$

$\dot{i}_L(0) = -m_1 \partial T_s$

$\dot{i}_L(T_s) = m_2 \partial T_s$

$\dot{i}_L(T_s) = \dot{i}_L(0) \left( \frac{-D}{D_s} \right)$

$\varepsilon_c(2T_s) = \varepsilon_c(0) \left( \frac{-C}{D_s} \right)^2$

$\varepsilon_c(nT_s) = \varepsilon_c(0) \left( \frac{-C}{D_s} \right)^n$

Stable only if $\frac{D}{D_s} < 1 \rightarrow [D < D_s]$
Final Value of Inductor Current

\[ i_L(T_s) - i_L(0) \left( -\frac{D}{D'} \right) \]

\[ i_L(2T_s) = i_L(T_s) \left( -\frac{D}{D'} \right) = i_L(0) \left( -\frac{D}{D'} \right)^2 \]

\[ i_L(nT_s) - i_L((n-1)T_s) \left( -\frac{D}{D'} \right) - i_L(0) \left( -\frac{D}{D'} \right)^n \]

\[ |i_L(nT_s)| \rightarrow \begin{cases} 
0 & \text{when } \left| -\frac{D}{D'} \right| < 1 \\
\infty & \text{when } \left| -\frac{D}{D'} \right| > 1 
\end{cases} \]

For stability: \( D < 0.5 \)

Example: Unstable operation for \( D=0.6 \)

\[ \alpha = -\frac{D}{D'} = \left( -\frac{0.6}{0.4} \right) = -1.5 \]
Example: Stable operation for $D=1/3$

$$\alpha = -\frac{D}{D'} = \left(-\frac{1/3}{2/3}\right) = -0.5$$

Stabilization Through Artificial Ramp

Now, transistor switches off when

$$i_s(dT_s) + i_R(dT_s) = i_c$$

or,

$$i_c(dT_s) = i_c - i_s(dT_s)$$
**Final Value of Inductor Current**

First subinterval:
\[ i_L(0) = -\alpha T_s \left( m_1 + m_a \right) \]

Second subinterval:
\[ i_L(T_s) = -\alpha T_s \left( m_a - m_2 \right) \]

Net change over one switching period:
\[ i_L(T_s) = i_L(0) \left( -\frac{m_2 - m_a}{m_1 + m_a} \right) \]

After \( n \) switching periods:
\[ i_L(nT_s) = i_L((n-1)T_s) \left( -\frac{m_2 - m_a}{m_1 + m_a} \right) = i_L(0) \left( -\frac{m_2 - m_a}{m_1 + m_a} \right)^n = i_L(0) \alpha^n \]

Characteristic value:
\[ \alpha = -\frac{m_2 - m_a}{m_1 + m_a} \]

\[ i_L(nT_s) \rightarrow \begin{cases} 0 & \text{when } |\alpha| < 1 \\ \infty & \text{when } |\alpha| > 1 \end{cases} \]

**Artificial Ramp: Additional Notes**

- For stability, require \(|\alpha| < 1\)
- Common choices:
  - \( m_a = 0.5 \ m_2 \)
  - \( m_a = m_2 \)
- Artificial ramp decreases sensitivity to noise
More Accurate Models

• The simple models of the previous section yield insight into the low-frequency behavior of CPM converters

• Unfortunately, they do not always predict everything that we need to know:
  − Line-to-output transfer function of the buck converter
  − Dynamics at frequencies approaching $f_s$

• More accurate model accounts for nonideal operation of current mode controller built-in feedback loop

• Converter duty-cycle-controlled model, plus block diagram that accurately models equations of current mode controller

• See Section 12.3 for additional info

Application to Experiment 4

• Complex switching controller

• Read the datasheet first
Internal Functional Model in LTSpice

- Accuracy/functionality not guaranteed
- Used for insight only

Workers only in 'forced pump'
- Not diode emulation
- Not burst mode
In-Circuit Simulation

.model mysw sw(Von=3 Voff=2 Ron=1 Roff = 1M)
.model myMOS VDMOS(Rg=1 Vto=4.5 Rd=14m Rs=18m Rb=17m Kp=30 Cgdm=5p Cgms=0.5m Cgs=2n Cj=0.05p 1s=88p)

Sim Results

Output Short

V(out)
A Tip: Debug Internal of Subcircuit