6.622 Power Electronics *Prof. David Perreault*

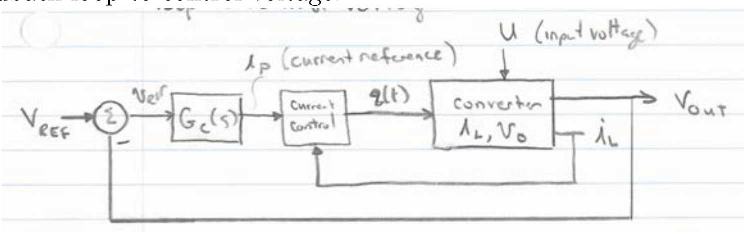
Lecture 27 - Control 4 (Current-Mode Control)

We have seen that duty-ratio control can yield (in Boost and other converters):

1. Nonlinear dynamics (vary w/ operating point)
2. RHP zero + lightly-damped poles

One can partially mitigate damping and operating-point variation with a damping leg. However, a still better strategy is "full-state feedback": control the converter based on both inductor current and output (capacitor) voltage.

In current-mode control we add an inner feedback loop to control inductor current, and use an outer feedback loop to control voltage.



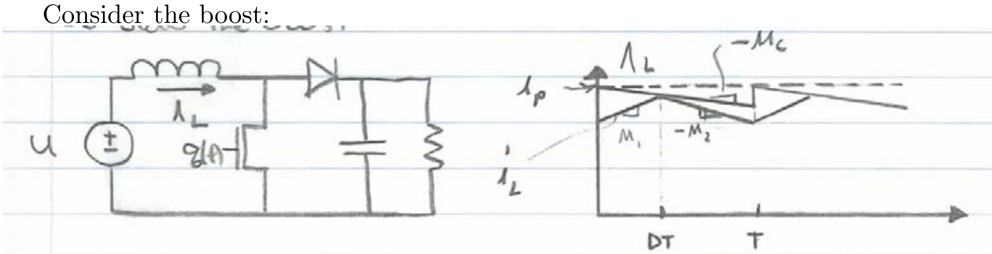
In peak current control we adjust the switching function q(t) such that

1. The switch turns on every T seconds
2. Switch turns off when the peak inductor current reaches (a simple function of) a reference value i_p

$\Rightarrow$ The value of i_p is set by the (outer) voltage control loop to regulate v_{out}
This gives

1. better controlled dynamics
2. cycle-by-cycle current limiting

Consider the boost:

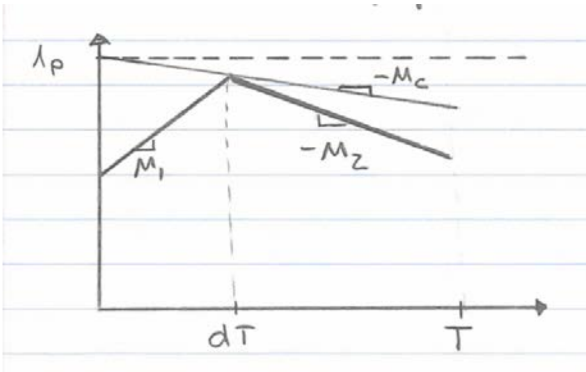


Boost
$\mu_1 = \frac{u}{L}$
$\mu_2 = \frac{v-u}{L}$

- Switch turns on at beginning of each cycle
- switch turns off when i_L reaches $i_p - \mu_c(t - nT)$
  - $\rightarrow$ will see reason for adjustment term μ_c shortly
  - $\rightarrow$ must sense i_L (or i_{sw}) but often do anyway for protection
- control output voltage by adjusting i_p

Consider system dynamics:

Simplest method: start with duty ratio equations, find (approximate) relation between d, i_p, $\overline{i_L}$



Look at a 1-cycle window and make a geometric approximation

$$\overline{i_L} \approx (i_p - \mu_c dT) - \frac{1}{T}[\frac{1}{2}\mu_1 d^2T^2 + \frac{1}{2}\mu_2(1-d)^2T^2]$$

$$\overline{i_L} \approx (i_p - \mu_c dT) - \frac{1}{2}\mu_1 d^2T + \frac{1}{2}\mu_2(1-d)^2T$$

Their equation can be used for various types of converters (but for different values of μ_1, μ_2

For a boost converter: $\mu_1 = \frac{\overline{u}}{L}, \mu_2 = \frac{\overline{v}-\overline{u}}{L}$

$$\therefore i_L = i_p - \mu_c dT - \frac{1}{2}\frac{\overline{u}T}{L}d^2 - \frac{(\overline{v}-\overline{u})T}{2L}(1-d)^2$$

linearize + solve for $\tilde{d}$:

$$(\text{Equation}*)\tilde{d} = \frac{1}{\mu_c T}(\tilde{i_p} - \tilde{i_L}) - \frac{(D^2 - D'^2)}{2L\mu_C}\tilde{u} - \frac{D'^2}{2L\mu_c}\tilde{v}$$

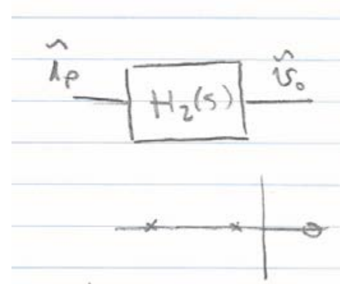
For boost converter

If we substitute Equation $*$ into the linearized state-spacemodel of the boost converter from before, we eliminated $\tilde{d}$ and have a new control variable $\tilde{i_p}$

From new model, we can get new linearized plant transfer function $H_2(s)$

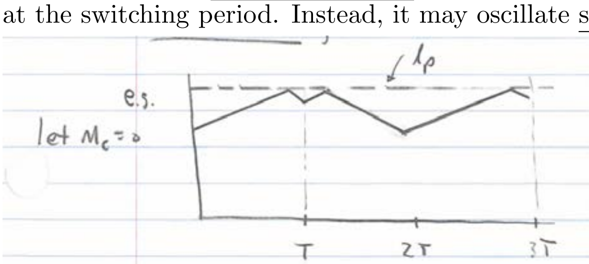
1. RHP zero
2. 2 LHP poles

   $\rightarrow$ low freq, 1 high freq, on real axis (depending on μ_c)



$\Rightarrow$ we can achieve much better control performance!

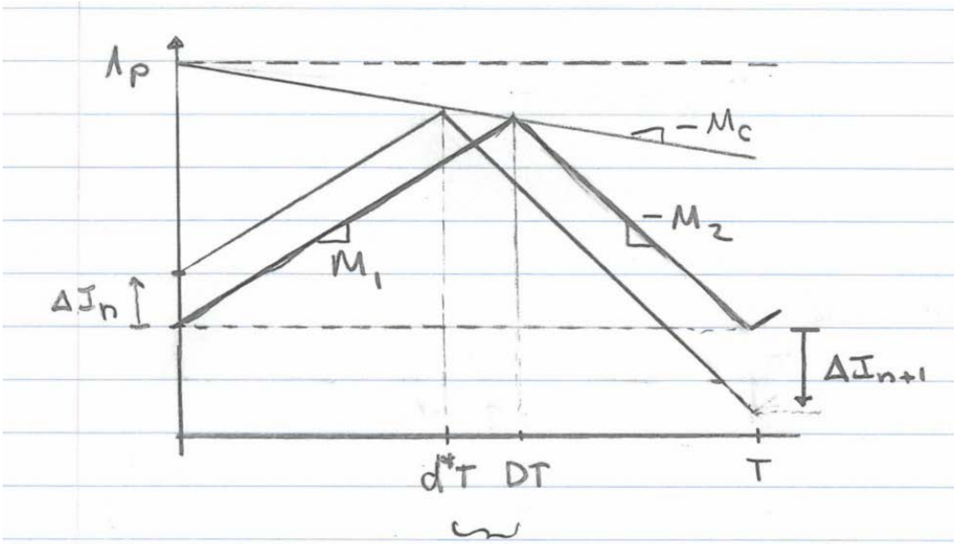
A challenge is Ripple Instability: under some conditions, the system will not settle to a single duty ratio at the switching period. Instead, it may oscillate subharmonically or chaotically!



Bad because

1. Low frequency ripple (below f_{sw})
2. Large ripple amplitude + spectrum
3. Control is "jittery"

A properly chosen "compensating ramp" (slope - $-\mu_c$) can fix this: $\Rightarrow$ let's analyze ripple dynamics (can't used averaged model)



Start w/ ΔI offset from steady state $\mu_1 D = \mu_2(1-D)$

Considering the parallel lines in this region

$$\Delta I_n = (\mu_1 + \mu_C)(D - d*)T$$

$$\Delta I_{n+1} = (\mu_C + \mu_2)(D - d*)T$$

$$\Delta I_{n+1} = -\frac{\mu_2 - \mu_C}{\mu_1 + \mu_C}\Delta I_n$$

$$\therefore \Delta I_n = (-\frac{\mu_2 - \mu_C}{\mu_1 + \mu_C})^n \Delta I_o$$

Unstable for

$$|\frac{\mu_2 - \mu_c}{\mu_1 + \mu_C}| > 1$$

@ $\mu_c = 0$ $|\frac{\mu_2}{\mu_1}| > 1 \overset{\text{boost}}{\rightarrow} |\frac{D}{1-D}| > 1 \therefore$ unstable for D > 0.5
$\Rightarrow$ choose μ_C to stabilize ripple dynamics ($\mu_2 = \mu_c$ deadbeat control)
But μ_C affects averaged dynamics (steeper μ_C looks more like duty ratio control) $\Rightarrow$ tradeoff.