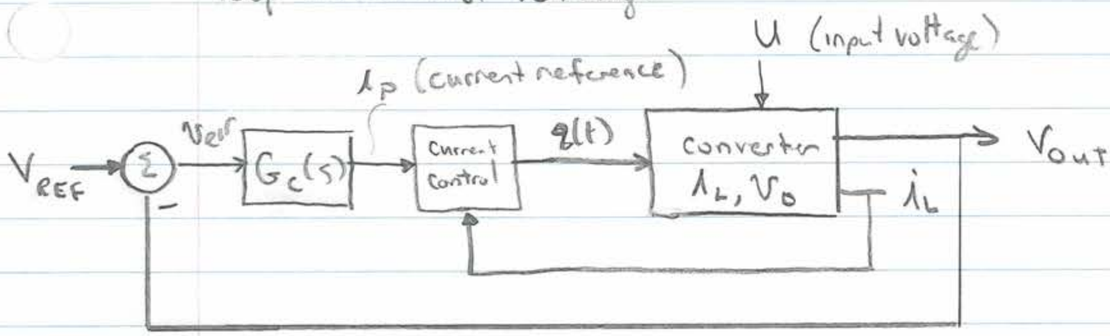
# 6.334 Lecture — Current-Mode Control

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 $i_p$.

⇒ 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

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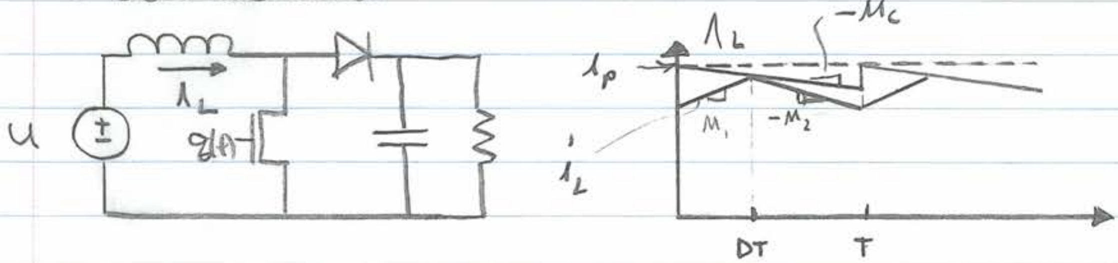
# Current-Mode Control

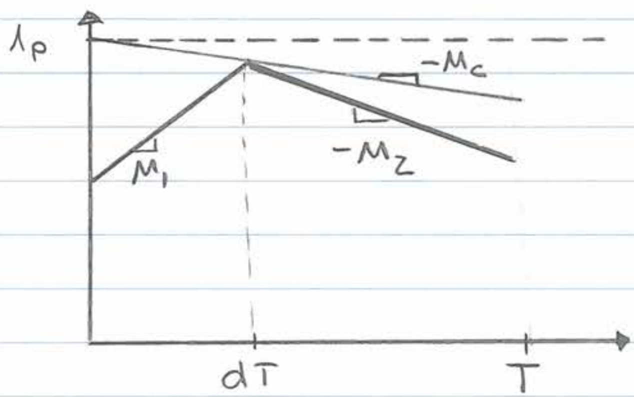
Consider the boost:




Boost
$M_1 = \frac{u}{L}$
$M_2 = \frac{v-u}{L}$

- Switch turns on at beginning of each cycle
- Switch turns off when $i_L$ reaches $i_p - M_c(t-nT)$
  - → will see reason for adjustment term $M_c$ shortly
  - → 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 eqns, find (approximate) relation between $d, i_p, \overline{i}_L$




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

$$\overline{i}_L \approx (i_p - M_c dT) - \frac{1}{T}\left[\frac{1}{2}M_1 d^2 T^2 + \frac{1}{2}M_2(1-d)^2 T^2\right]$$

$$\therefore \boxed{\overline{i}_L \approx i_p - M_c dT - \frac{1}{2}M_1 d^2 T - \frac{1}{2}M_2(1-d)^2 T}$$

This equation can be used for various types of converters (but for different values of $M_1, M_2$

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for a <u>boost converter</u>: $M_1 = \frac{\bar{u}}{L}$, $M_2 = \frac{\bar{v} - \bar{u}}{L}$

$$\therefore \lambda_L = \lambda_p - M_c d T - \frac{1}{2}\frac{\bar{u}T}{L} d^2 - \frac{(\bar{v} - \bar{u})T}{2L}(1-d)^2$$

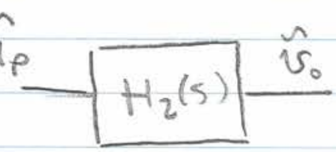
Linearize + solve for $\hat{d}$:

$$\star \quad \boxed{\hat{d} = \frac{1}{M_c T}(\hat{\lambda}_p - \hat{\lambda}_L) - \frac{(D^2 - D'^2)}{2LM_c}\hat{u} - \frac{D'^2}{2LM_c}\hat{v}}$$

for boost converter

If we substitute ($\star$) into the linearized state-space model of the boost converter from before, we eliminate $\hat{d}$ and have a <u>new</u> control variable $\hat{\lambda}_p$

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




⇒ 1.) RHP zero

2.) 2 LHP poles

→ low freq, 1 high freq on real axis (depending on $M_c$)

⇒ we can achieve much better control performance!

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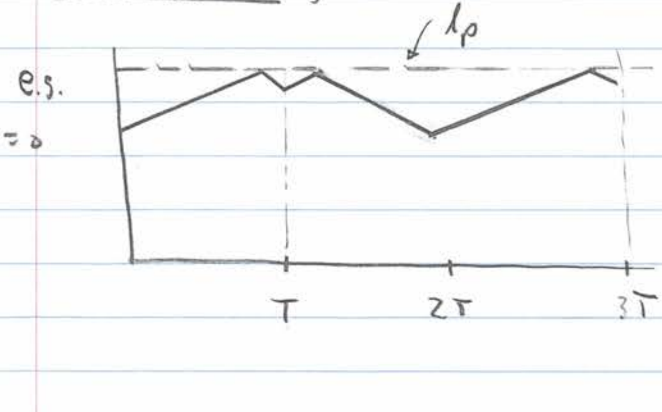
A challenge is <u>Ripple Instability</u>: under some conditions, the system will not settle to a single duty ratio at the switching period. Instead, it may oscillate <u>subharmonically</u> or <u>chaotically</u>!

let $M_c = 0$




Bad because

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

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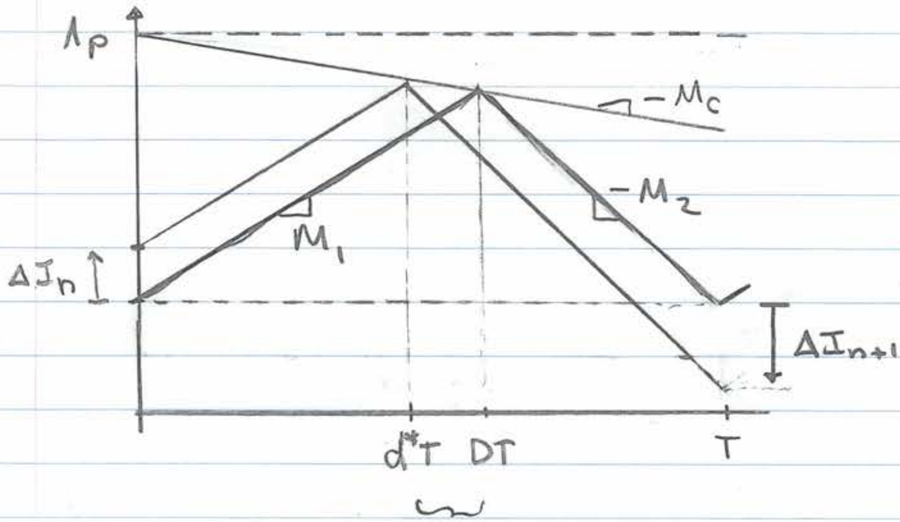
A properly chosen "Compensating ramp" (slope $= -M_c$) can fix this:
$\Rightarrow$ lets analyze ripple dynamics (can't use averaged model)




START w/ $\Delta I$ offset from STEADY STATE
$M_1 D = M_2(1-D)$

Considering the parallel lines in this region

$$\begin{cases} \Delta I_n = (M_1 + M_c)(D - d^*)T \\ \Delta I_{n+1} = (M_c - M_2)(D - d^*)T \end{cases}$$

$$\therefore \quad \Delta I_{n+1} = -\frac{M_2 - M_c}{M_1 + M_c} \Delta I_n$$

$$\therefore \quad \Delta I_n = \left(-\frac{M_2 - M_c}{M_1 + M_c}\right)^n \Delta I_0$$

$\boxed{\text{unstable for } \left|\frac{M_2 - M_c}{M_1 + M_c}\right| > 1}$

@ $M_c = 0$ $\quad \left|\frac{M_2}{M_1}\right| > 1 \xrightarrow{\text{BOOST}} \left|\frac{D}{1-D}\right| > 1 \quad \therefore$ unstable for $D > 0.5$

$\Rightarrow$ Choose $M_c$ to stabilize ripple dynamics ($M_2 = M_c$ deadbeat control)

But $M_c$ affects averaged dynamics (steeper $M_c$ looks more like duty ratio control) $\Rightarrow$ Tradeoff.