Lecture 10: Isolated Converters II & DCM Introduction

ECE 481: Power Electronics
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Announcements

- Midterm exam due Thursday, start of class
- Hw #3 returned today
- Missing HW assignments

3. [30 pts] Design of a Boost Converter

The boost converter in Fig. 3 connects a lead-acid battery to a 48 V DC bus. The converter input is the battery voltage, which has characteristics:

Maximum V_{batt}: 15 V
Minimum V_{batt}: 10 V

The ~~maximum~~ output power is 100 W, and the switching frequency is 100 kHz. The inductance L is chosen to be 10 μH, and the capacitance may be assumed to be very large for parts (a)-(c).

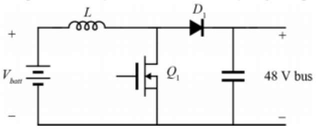

Figure 3: Boost converter

a) **[10 pts]** A number of parts are available for both the MOSFET and diode. Important characteristics of each device are shown in Tables I and II. The rated maximum currents and voltages are the maximum *instantaneous* values which the devices can handle.

TABLE I: MOSFET DEVICES

Device	Rated max V_{DS}	Rated max I_{DS}	R_{on}
I	40 V	30 A	5 mΩ
II	100 V	15 A	15 mΩ
III	100 V	20 A	50 mΩ
IV	150 V	15 A	30 mΩ
V	150 V	20 A	75 mΩ

TABLE II: DIODE DEVICES

Device	Rated max V_R	Rated max I_F	V_F
VI	40 V	20 A	0.5 V
VII	100 V	15 A	1.0 V
VIII	150 V	5 A	0.7 V
IX	200 V	20 A	1.5 V

Select one MOSFET and one diode which will work best in this converter. Explain why you

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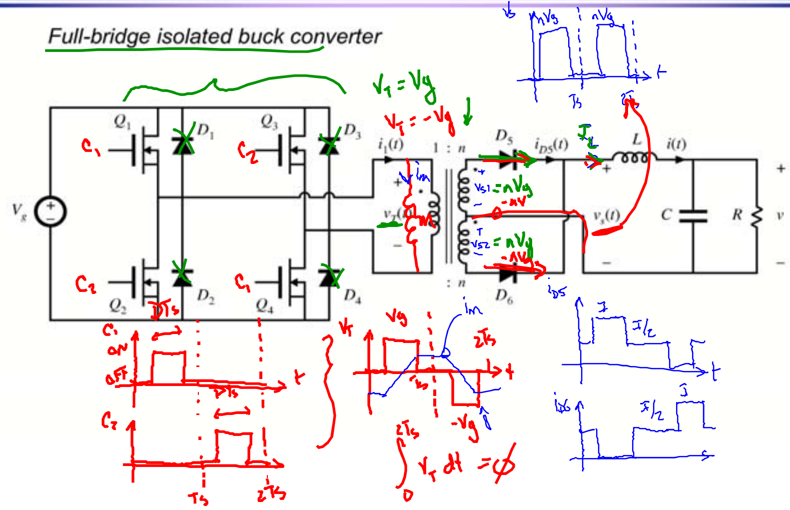
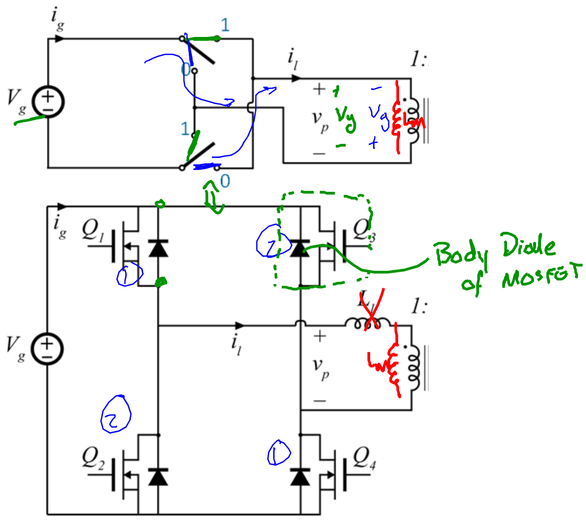
6.3.1. Full-bridge and half-bridge isolated buck converters

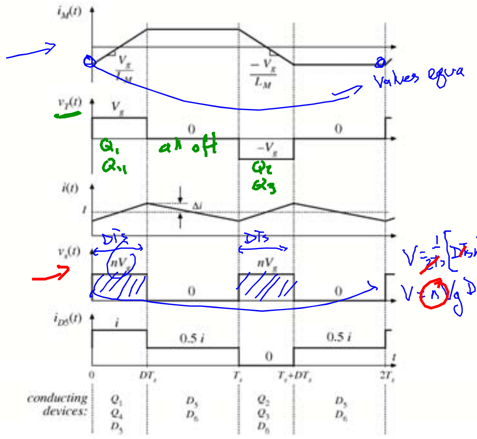
Full-bridge isolated buck converter



Full Bridge Switch Structure



Full-bridge: waveforms



$V = \frac{1}{\cancel{2}T_s}\left[D\cancel{T_s} n V_g \cancel{2}\right]$

$V = nV_g D$

- During first switching period: transistors Q_1 and Q_4 conduct for time DT_s, applying volt-seconds $V_g\,DT_s$ to primary winding
- During next switching period: transistors Q_2 and Q_3 conduct for time DT_s, applying volt-seconds $-V_g\,DT_s$ to primary winding
- Transformer volt-second balance is obtained over two switching periods
- Effect of nonidealities?

Effect of nonidealities on transformer volt-second balance

Volt-seconds applied to primary winding during first switching period:

Different r_{on} of Q_1 & Q_4

$(V_g - (Q_1 \text{ and } Q_4 \text{ forward voltage drops}))(Q_1 \text{ and } Q_4 \text{ conduction time})$

Volt-seconds applied to primary winding during next switching period:

$-(V_g - (Q_2 \text{ and } Q_3 \text{ forward voltage drops}))(Q_2 \text{ and } Q_3 \text{ conduction time})$

These volt-seconds never add to *exactly* zero.

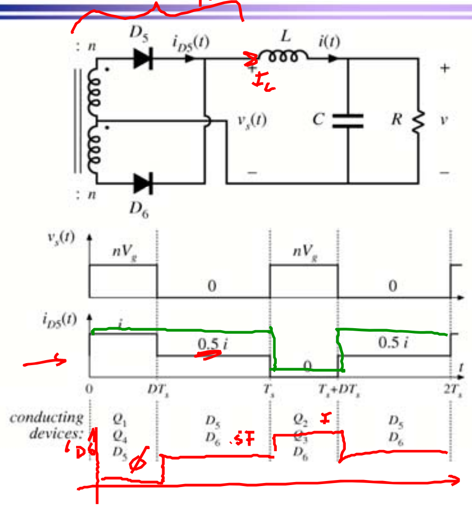
Net volt-seconds are applied to primary winding

Magnetizing current slowly increases in magnitude

Saturation can be prevented by placing a capacitor in series with primary, or by use of current programmed mode (Chapter 12)

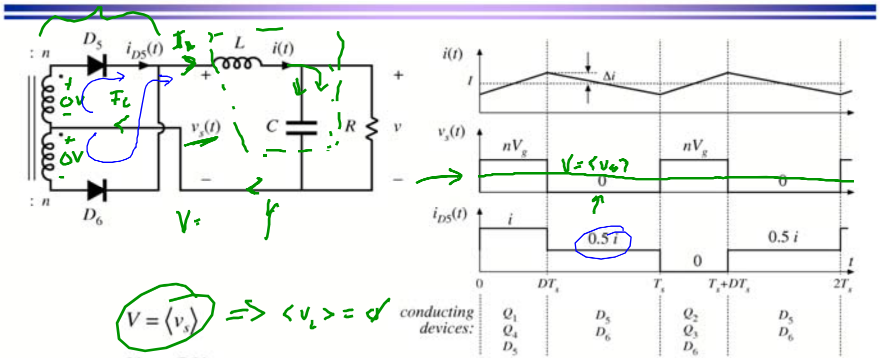
Operation of secondary-side diodes

Center-tapped rectifier



- During second (D') subinterval, both secondary-side diodes conduct
- Output filter inductor current divides approximately equally between diodes
- Secondary amp-turns add to approximately zero
- Essentially no net magnetization of transformer core by secondary winding currents

Volt-second balance on output filter inductor

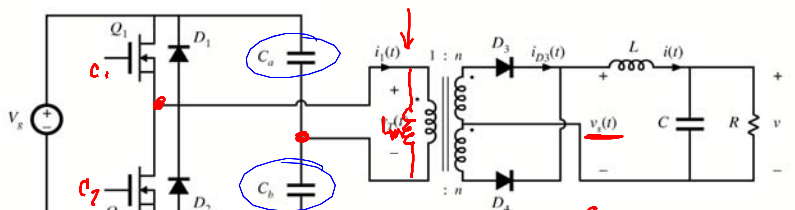


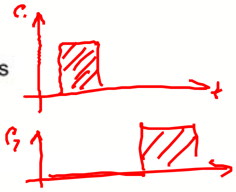
$\boxed{V = \langle v_s \rangle} \Rightarrow \langle v_L \rangle = 0$

$V = nDV_g$

$M(D) = \boxed{nD}$ buck converter with turns ratio

Half-bridge isolated buck converter

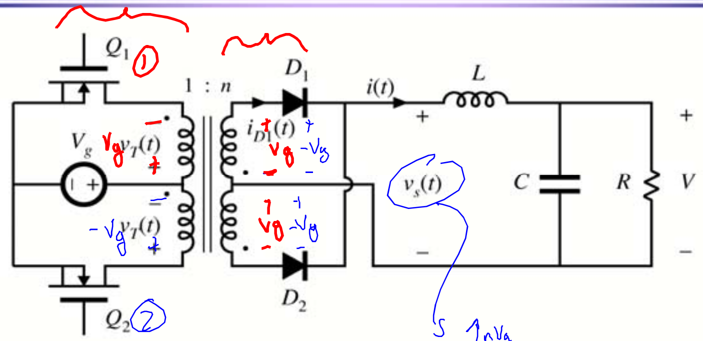




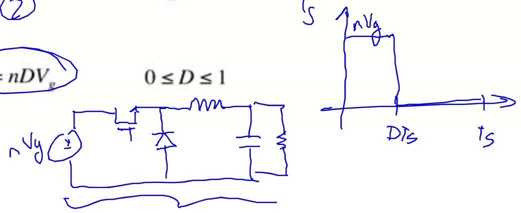
- Replace transistors Q_3 and Q_4 with large capacitors
- Voltage at capacitor centerpoint is $0.5V_g$
- $v_s(t)$ is reduced by a factor of two
- $M = 0.5\, nD$

$$\langle v_{Lm} \rangle = D(V_g - V_{Cb}) + D(0 - V_{Cb})$$

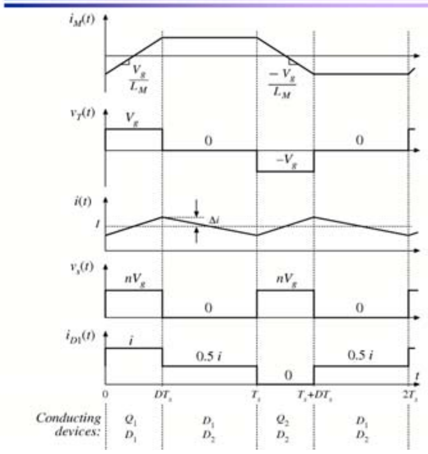
6.3.3. Push-pull isolated buck converter



$V = nDV_g$ $\qquad 0 \le D \le 1$

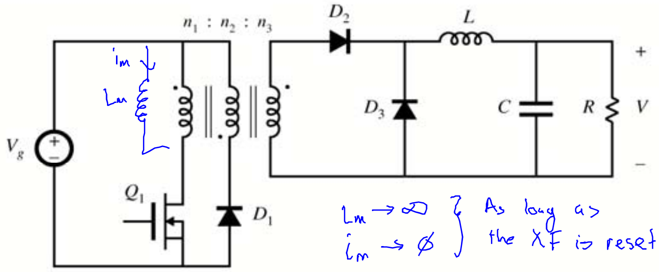


Waveforms: push-pull



- Used with low-voltage inputs
- Secondary-side circuit identical to full bridge
- As in full bridge, transformer volt-second balance is obtained over two switching periods
- Effect of nonidealities on transformer volt-second balance?
- Current programmed control can be used to mitigate transformer saturation problems. Duty cycle control not recommended.

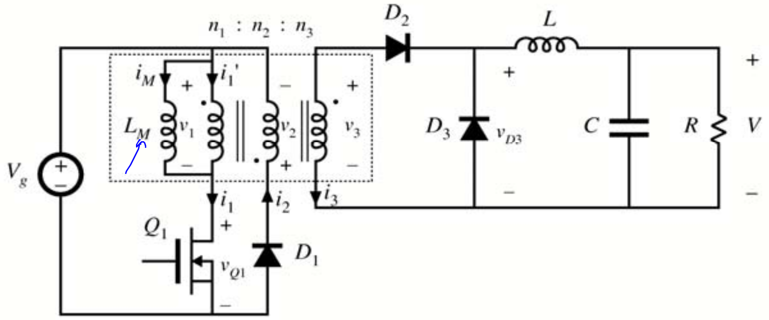
6.3.2. Forward converter



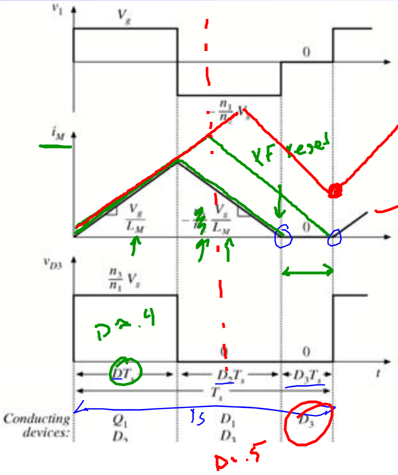
$L_m \to \infty$
$i_m \to 0$ } As long as the XF is reset

- Buck-derived transformer-isolated converter
- Single-transistor and two-transistor versions
- Maximum duty cycle is limited
- Transformer is reset while transistor is off

Forward converter
with transformer equivalent circuit



Forward converter: waveforms



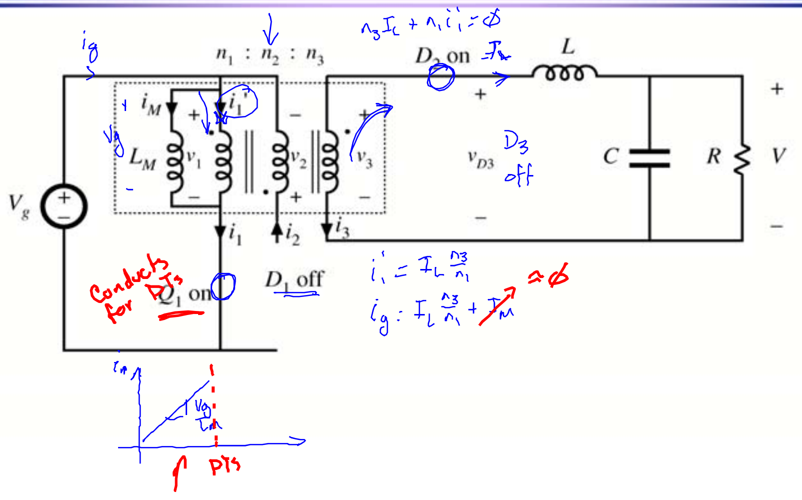
- Magnetizing current, in conjunction with diode D_1, operates in discontinuous conduction mode
- Output filter inductor, in conjunction with diode D_3, may operate in either CCM or DCM

saturation

if $n_1 = n_2$

$D < .5$ for transformer reset

Subinterval 1: transistor conducts

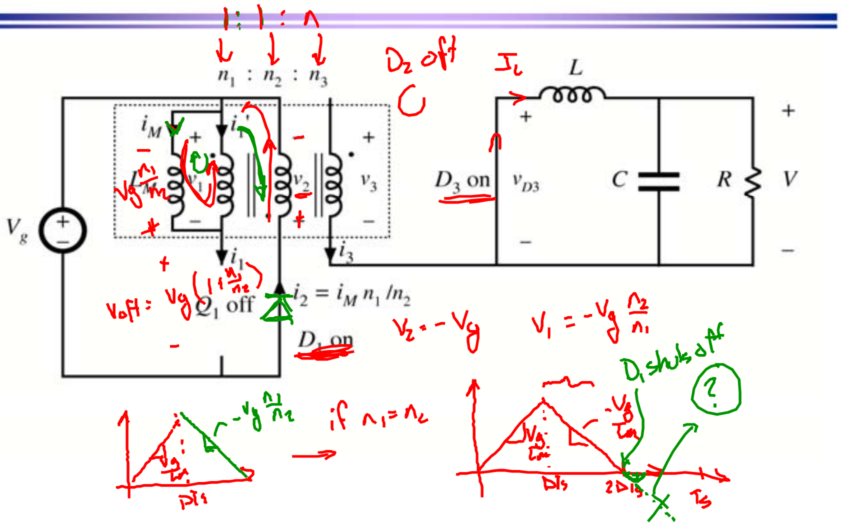


$n_3 I_L + n_1 i_1' = 0$

$i_1' = I_L \frac{n_3}{n_1}$

$i_g = I_L \frac{n_3}{n_1} + \cancel{I_M} \approx 0$

Subinterval 2: transformer reset

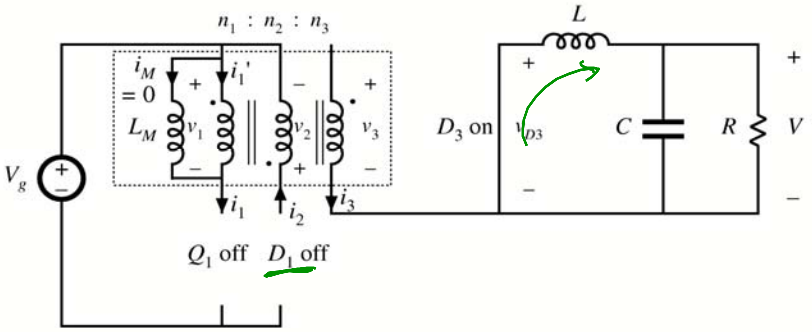


$V_{off} = V_g\left(1+\frac{n_1}{n_2}\right)$

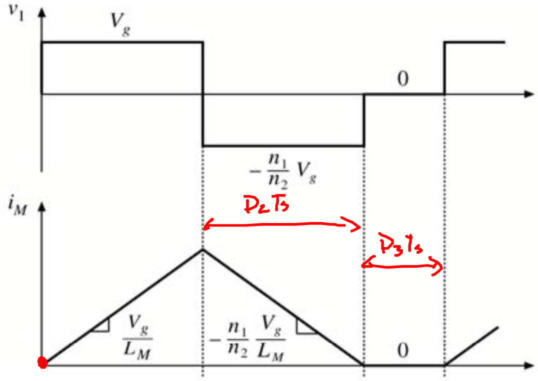
$V_2 = -V_g$

$V_1 = -V_g \frac{n_2}{n_1}$

Subinterval 3



Magnetizing inductance volt-second balance



$$\langle v_1 \rangle = D\left(V_g\right) + \underline{D_2\left(-V_g n_1/n_2\right)} + D_3\left(0\right) = 0$$

Transformer reset

From magnetizing current volt-second balance:

$$\langle v_1 \rangle = D\left(V_g\right) + D_2\left(-V_g n_1/n_2\right) + D_3\left(0\right) = 0$$

Solve for D_2:

$$\underline{D_2 = \frac{n_2}{n_1}} D$$

D_3 cannot be negative. But $D_3 = 1 - D - D_2$. Hence

$$D_3 = 1 - D - D_2 \geq 0$$

$$D_3 = 1 - D\left(1 + \frac{n_2}{n_1}\right) \geq 0$$

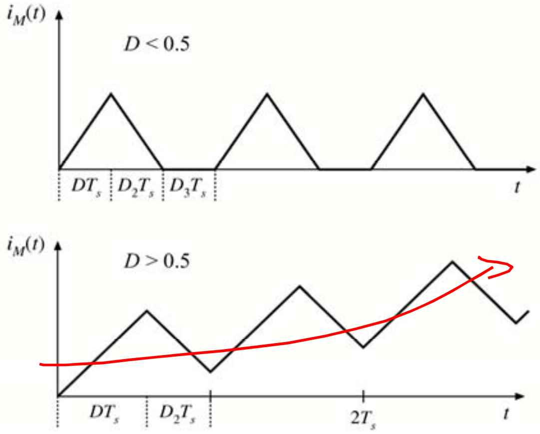
Solve for D

$$D \leq \frac{1}{1 + \frac{n_2}{n_1}}$$

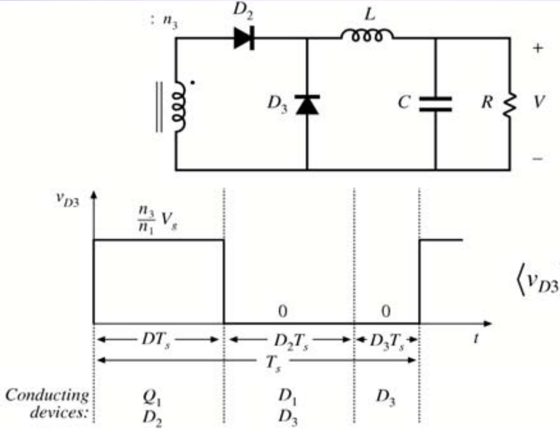
$\underline{\text{for } n_1 = n_2\text{:} \quad D \leq \frac{1}{2}}$

What happens when $D > 0.5$

magnetizing current waveforms,
for $\boxed{n_1 = n_2}$



Conversion ratio $M(D)$



$$\langle v_{D3} \rangle = \underline{V = \frac{n_3}{n_1} DV_g}$$

Maximum duty cycle vs. transistor voltage stress

Maximum duty cycle limited to

$$D \le \frac{1}{1 + \frac{n_2}{n_1}}$$

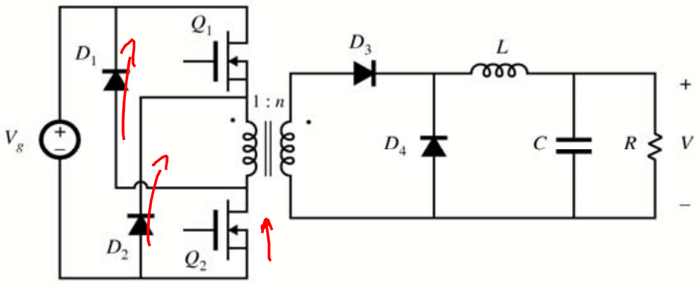
which can be increased by increasing the turns ratio n_2 / n_1. But this increases the peak transistor voltage:

$$\max\left(v_{Q1}\right) = V_g\left(1 + \frac{n_1}{n_2}\right)$$

For $n_1 = n_2$

$D \le \frac{1}{2}$ and $\max(v_{Q1}) = 2V_g$

The two-transistor forward converter



$V = nDV_g$ $D \le \frac{1}{2}$ $\max(v_{Q1}) = \max(v_{Q2}) = V_g$

Chapter 5. The Discontinuous Conduction Mode

5.1. Origin of the discontinuous conduction mode, and mode boundary

5.2. Analysis of the conversion ratio $M(D,K)$

5.3. Boost converter example

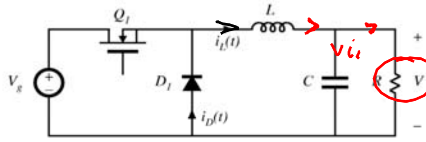
5.4. Summary of results and key points

Introduction to Discontinuous Conduction Mode (DCM)

- Occurs because switching ripple in inductor current or capacitor voltage causes polarity of applied switch current or voltage to reverse, such that the current- or voltage-unidirectional assumptions made in realizing the switch are violated.
- Commonly occurs in dc-dc converters and rectifiers, having single-quadrant switches. May also occur in converters having two-quadrant switches.
- Typical example: dc-dc converter operating at light load (small load current). Sometimes, dc-dc converters and rectifiers are purposely designed to operate in DCM at all loads.
- Properties of converters change radically when DCM is entered:
  M becomes load-dependent
  Output impedance is increased
  Dynamics are altered
  Control of output voltage may be lost when load is removed

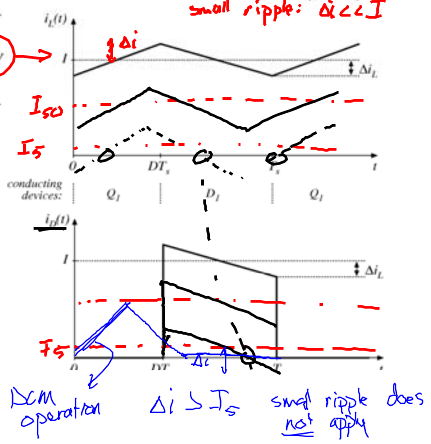
5.1. Origin of the discontinuous conduction mode, and mode boundary

Buck converter example, with single-quadrant switches



continuous conduction mode (CCM)

small ripple: $\Delta i << I$



DCM operation

$\Delta i > I_S$

small ripple does not apply

Minimum diode current is $(I - \Delta i_L)$

Dc component $I = V/R$

Current ripple is

$$\Delta i_L = \frac{(V_g - V)}{2L} DT_s = \boxed{\frac{V_g DD'T_s}{2L}}$$

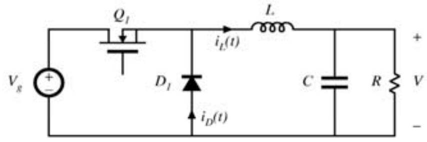
Note that I depends on load, but Δi_L does not.

if $\langle i_c \rangle = 0$

$I_L = \frac{V}{R}$

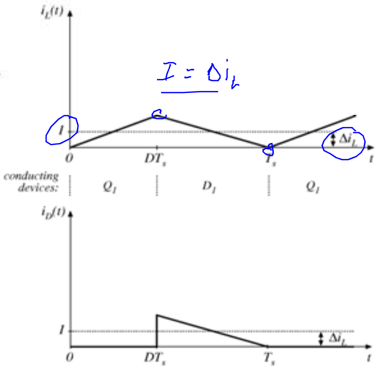
Reduction of load current

Increase R, until $I = \Delta i_L$



"Continuous Cond. Mode"

CCM-DCM boundary



Minimum diode current is $(I - \Delta i_L)$

Dc component $I = V/R$

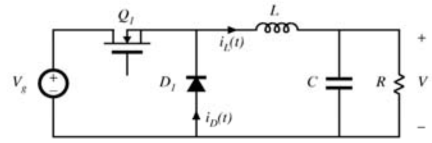
Current ripple is

$$\Delta i_L = \frac{(V_g - V)}{2L} DT_s = \frac{V_g DD'T_s}{2L}$$

Note that I depends on load, but Δi_L does not.

Further reduce load current

Increase R some more, such that $I < \Delta i_L$

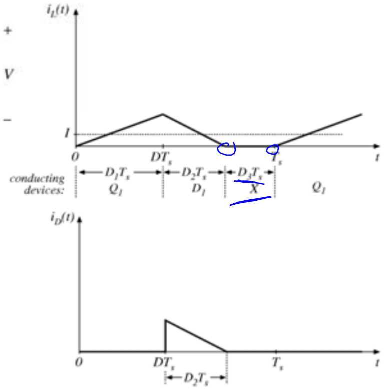


Minimum diode current is $(I - \Delta i_L)$
Dc component $I = V/R$
Current ripple is

$$\Delta i_L = \frac{(V_g - V)}{2L} DT_s = \frac{V_g DD'T_s}{2L}$$

Note that I depends on load, but Δi_L does not.
The load current continues to be positive and non-zero.

Discontinuous conduction mode



Mode boundary

Boundary
@ $I = \Delta i_L$

$I > \Delta i_L$ *for CCM*
$I < \Delta i_L$ *for DCM*

Insert buck converter expressions for I and Δi_L :

$V = DV_g$
$I_L = \frac{V}{R} = \frac{DV_g}{R}$

$$\frac{DV_g}{R} < \frac{DD'T_sV_g}{2L}$$

$I_L < \Delta i_L$

Simplify:

$$\boxed{\frac{2L}{RT_s} < D'}$$

for buck
$K_{crit}(D) = D'$

This expression is of the form

$$\boxed{K < K_{crit}(D)} \quad for\ DCM$$

where $\boxed{K = \frac{2L}{RT_s}}$ and $K_{crit}(D) = D'$

unitless parameter

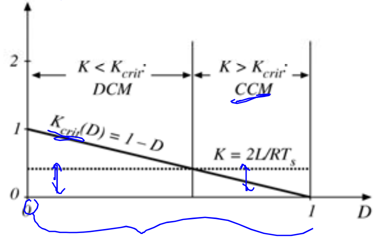
as $L\downarrow$ $\Delta i_L \uparrow$
↳ closer to DCM
as $T_s\downarrow$ $\Delta i_L \downarrow$
↳ farther from DCM
as $R\downarrow$ $I_L \uparrow$
↳ farther from DCM

K and K_{crit} vs. D

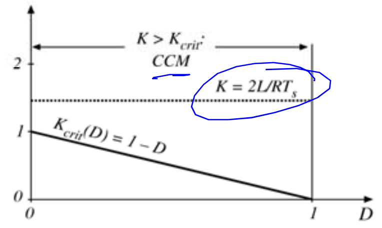
$K = \frac{2L}{RT_s}$

↑

for $K < 1$:



for $K > 1$:



Critical load resistance R_{crit}

Solve K_{crit} equation for load resistance R:

$R < R_{crit}(D)$ *for CCM*

→ $R > R_{crit}(D)$ *for DCM*

where $R_{crit}(D) = \frac{2L}{D'T_s}$

Summary: mode boundary

$$K > K_{crit}(D) \quad or \quad R < R_{crit}(D) \quad for\ CCM$$
$$K < K_{crit}(D) \quad or \quad R > R_{crit}(D) \quad for\ DCM$$

Table 5.1. CCM-DCM mode boundaries for the buck, boost, and buck-boost converters

Converter	$K_{crit}(D)$	$\max_{0 \le D \le 1}(K_{crit})$	$R_{crit}(D)$	$\min_{0 \le D \le 1}(R_{crit})$
Buck	$(1-D)$	1	$\frac{2L}{(1-D)T_s}$	$2\frac{L}{T_s}$
Boost	$D(1-D)^2$	$\frac{4}{27}$	$\frac{2L}{D(1-D)^2 T_s}$	$\frac{27}{2}\frac{L}{T_s}$
Buck-boost	$(1-D)^2$	1	$\frac{2L}{(1-D)^2 T_s}$	$2\frac{L}{T_s}$

5.2. Analysis of the conversion ratio *M(D,K)*

Analysis techniques for the discontinuous conduction mode:

Inductor volt-second balance

$$\langle v_L \rangle = \frac{1}{T_s}\int_0^{T_s} v_L(t)\,dt = 0$$

Capacitor charge balance

$$\langle i_C \rangle = \frac{1}{T_s}\int_0^{T_s} i_C(t)\,dt = 0$$

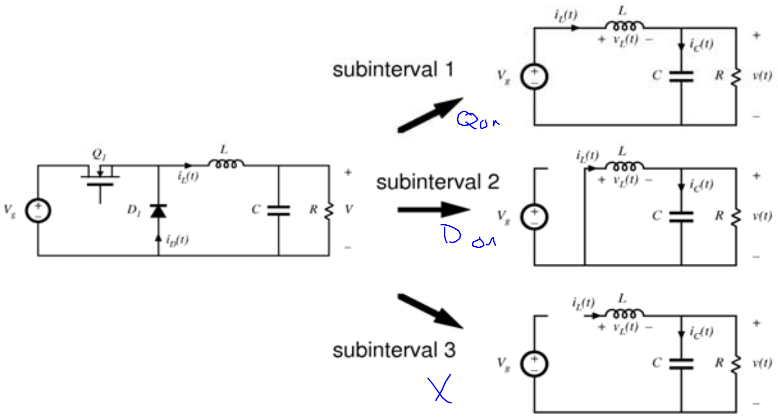
Small ripple approximation sometimes applies:

$\underline{v(t) \approx V}$ because $\Delta v << V$ Assuming operating as DC voltage regulator

No small ripple on i_L L — $\boxed{i(t) \approx I}$ is a poor approximation when $\Delta i > I$

Converter steady-state equations obtained via charge balance on each capacitor and volt-second balance on each inductor. Use care in applying small ripple approximation.

Example: Analysis of DCM buck converter $M(D,K)$



subinterval 1

Qon

subinterval 2

D on

subinterval 3

X