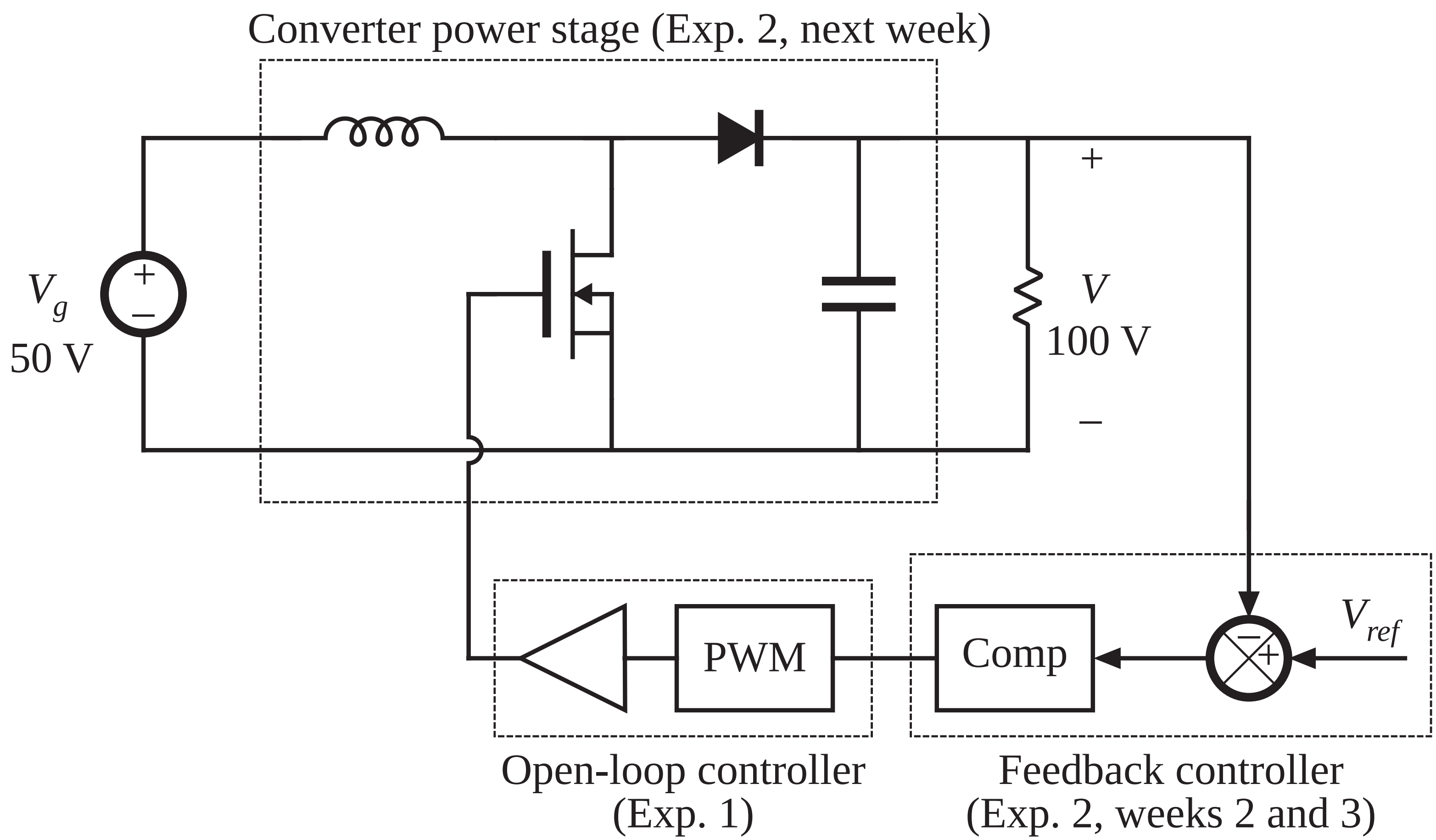
Experiment 2
Boost Switching DC-DC Power Converter

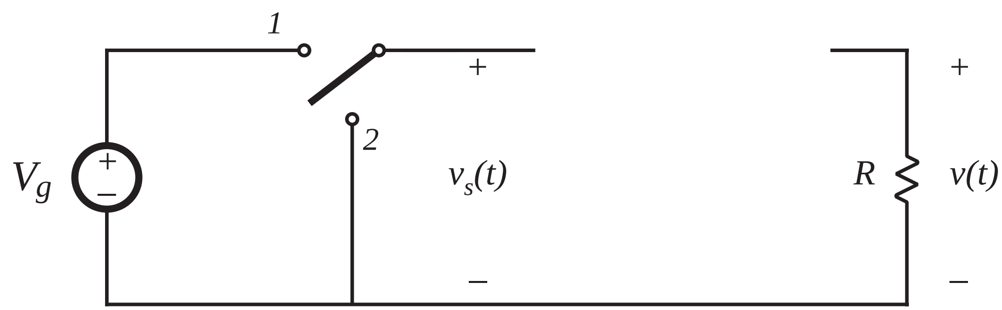


Today's lecture
Objectives

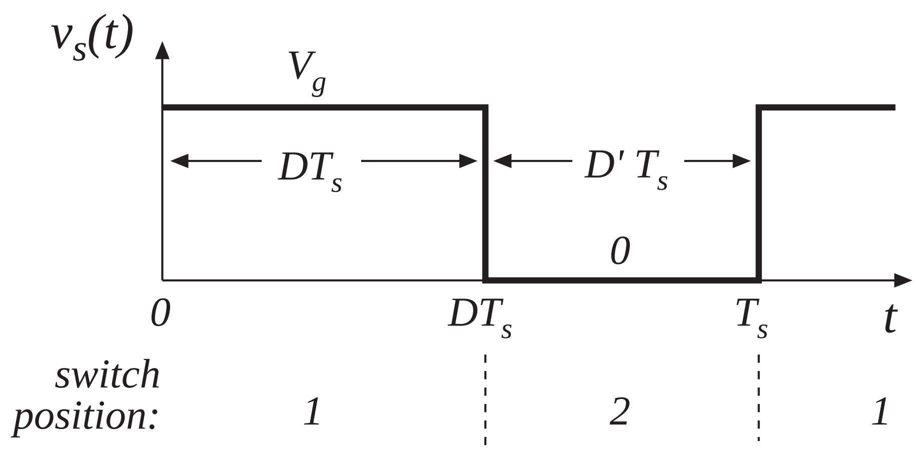
- Understanding the boost converter
  - Waveforms
  - Operation of transistor and diode
- Analysis of boost circuit
  - Basic approach and approximations
  - Design equations
- Inductor design
  - Design constraints
  - Design procedure

Last week's lecture:
Buck converter

SPDT switch changes dc component



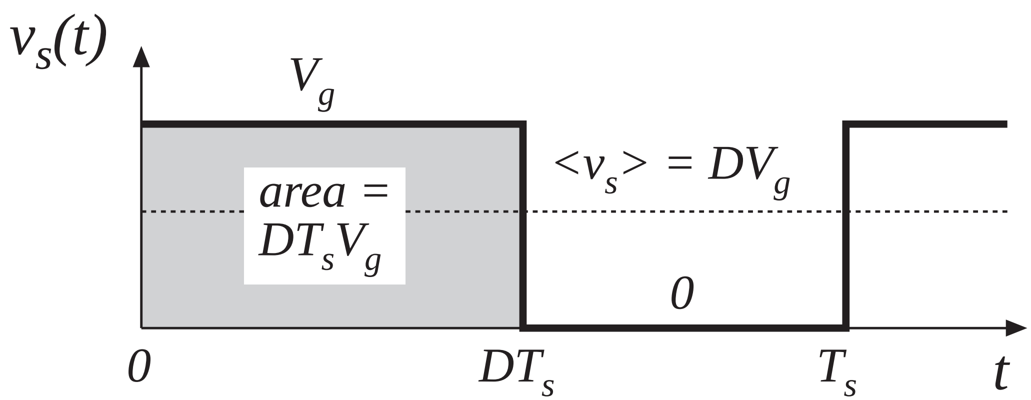
Switch output voltage waveform



Duty cycle *D:*
$0 \leq D \leq 1$

complement D':
$D' = 1 - D$

DC component of switch output voltage

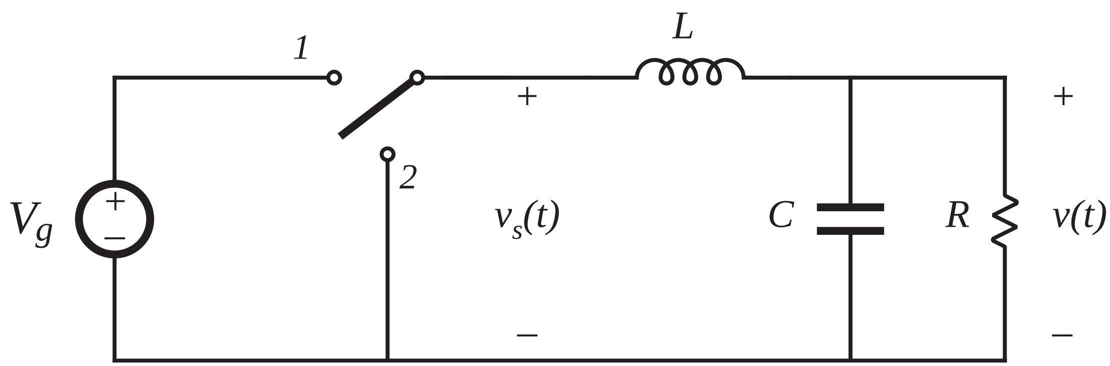


Fourier analysis: Dc component = average value

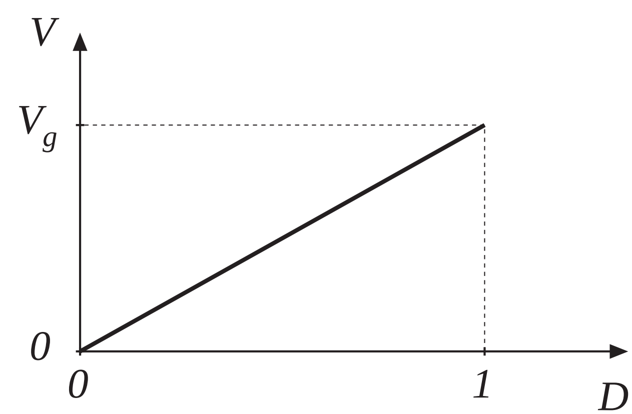
$$\langle v_s \rangle = \frac{1}{T_s}\int_0^{T_s} v_s(t)\, dt$$

$$\langle v_s \rangle = \frac{1}{T_s}\,(DT_sV_g) = DV_g$$

Insertion of low-pass filter to remove switching harmonics and pass only dc component



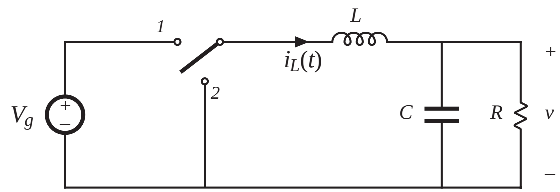
$$v \approx \langle v_s \rangle = DV_g$$

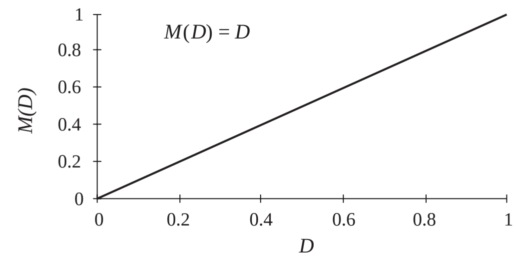


Three basic dc-dc converters

Buck

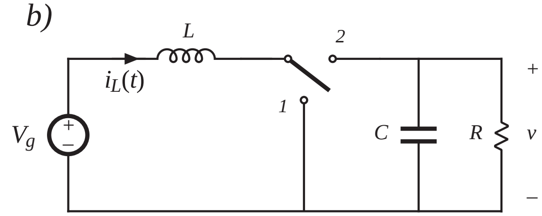
a)

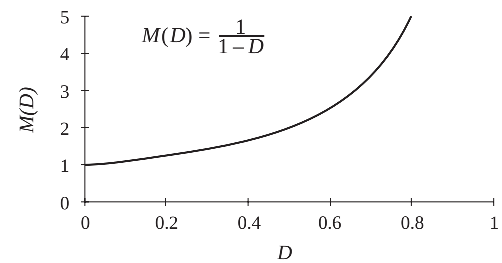




Boost

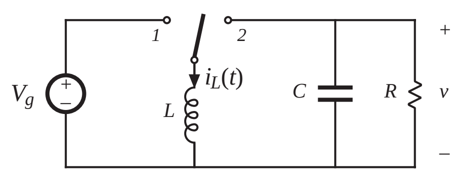
b)

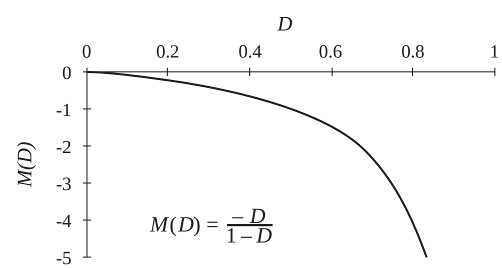




Buck-boost

c)

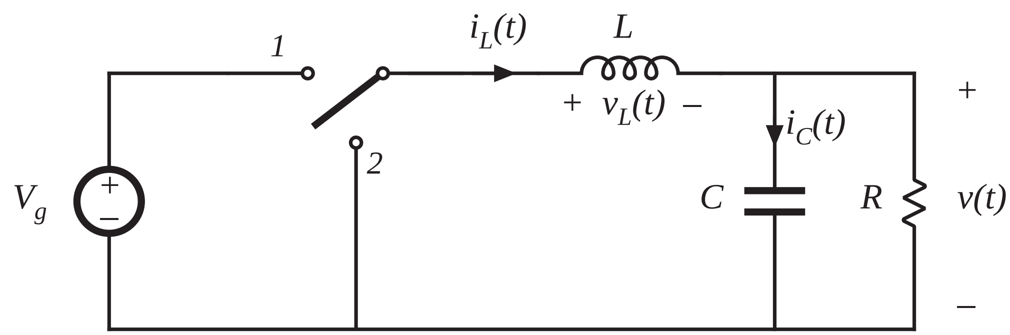




Analysis: Inductor volt-second balance, capacitor charge balance, and the small ripple approximation

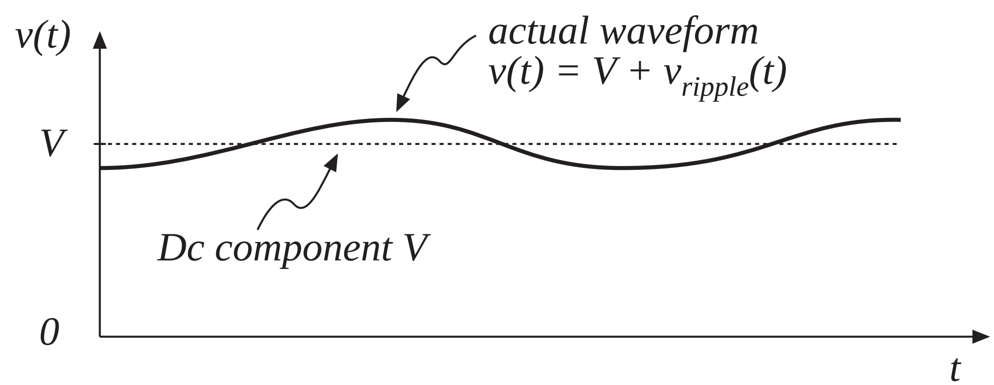
Actual output voltage waveform, buck converter

Buck converter containing practical low-pass filter



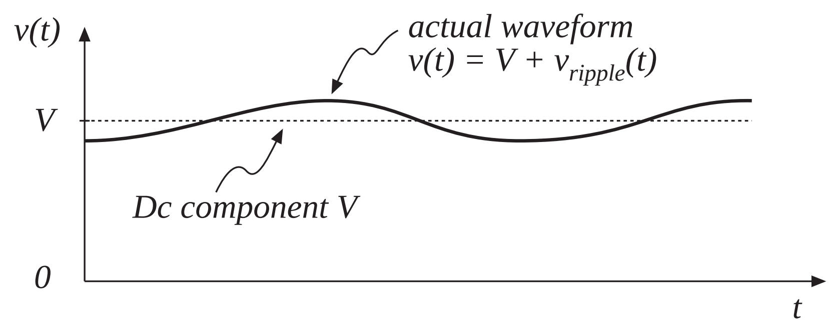
Actual output voltage waveform

$$v(t) = V + v_{ripple}(t)$$



The small ripple approximation

$$v(t) = V + v_{ripple}(t)$$



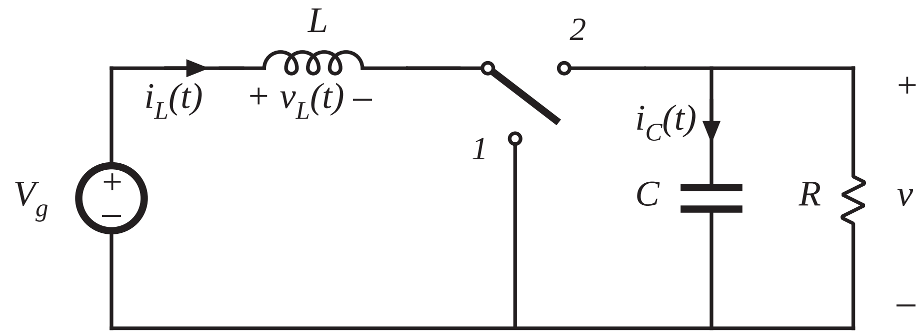
In a well-designed converter, the output voltage ripple is small. Hence, the waveforms can be easily determined by ignoring the ripple:

$$\left\| v_{ripple} \right\| << V$$

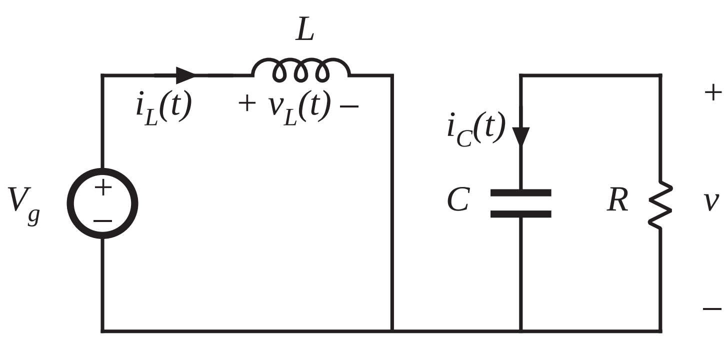
$$v(t) \approx V$$

Boost converter analysis: Inductor current waveform

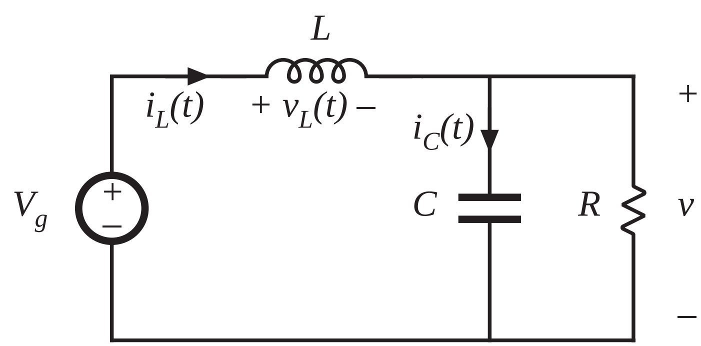
original converter



switch in position 1



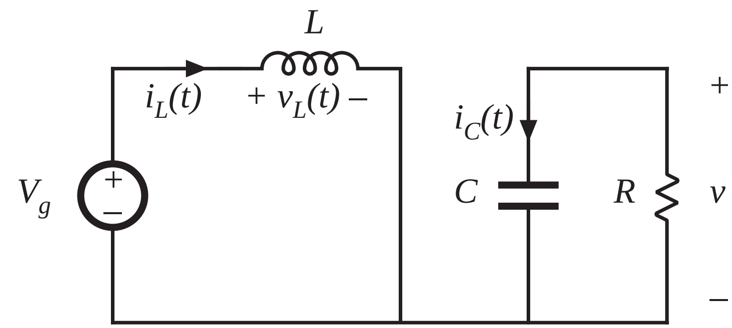
switch in position 2



Subinterval 1: switch in position 1
Let's find the inductor voltage and current

Inductor voltage and capacitor current

$$v_L = V_g$$
$$i_C = -v / R$$



Small ripple approximation:

$$v_L = V_g$$
$$i_C = -V / R$$

Knowing the inductor voltage, we can now find the inductor current via

$$v_L(t) = L \frac{di_L(t)}{dt}$$

Solve for the slope:

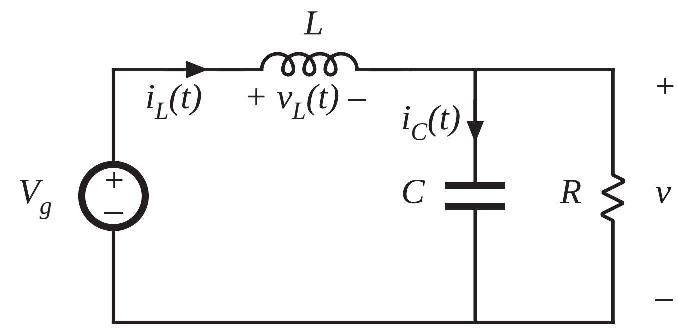
$$\frac{di_L(t)}{dt} = \frac{v_L(t)}{L} = \frac{V_g}{L}$$

⇒ *The inductor current increases with constant slope Vg/L*

Subinterval 2: switch in position 2

Inductor voltage and capacitor current

$$v_L = V_g - v$$
$$i_C = i_L - v / R$$



Small ripple approximation:

$$v_L = V_g - V$$
$$i_C = I - V / R$$

Knowing the inductor voltage, we can again find the inductor current via

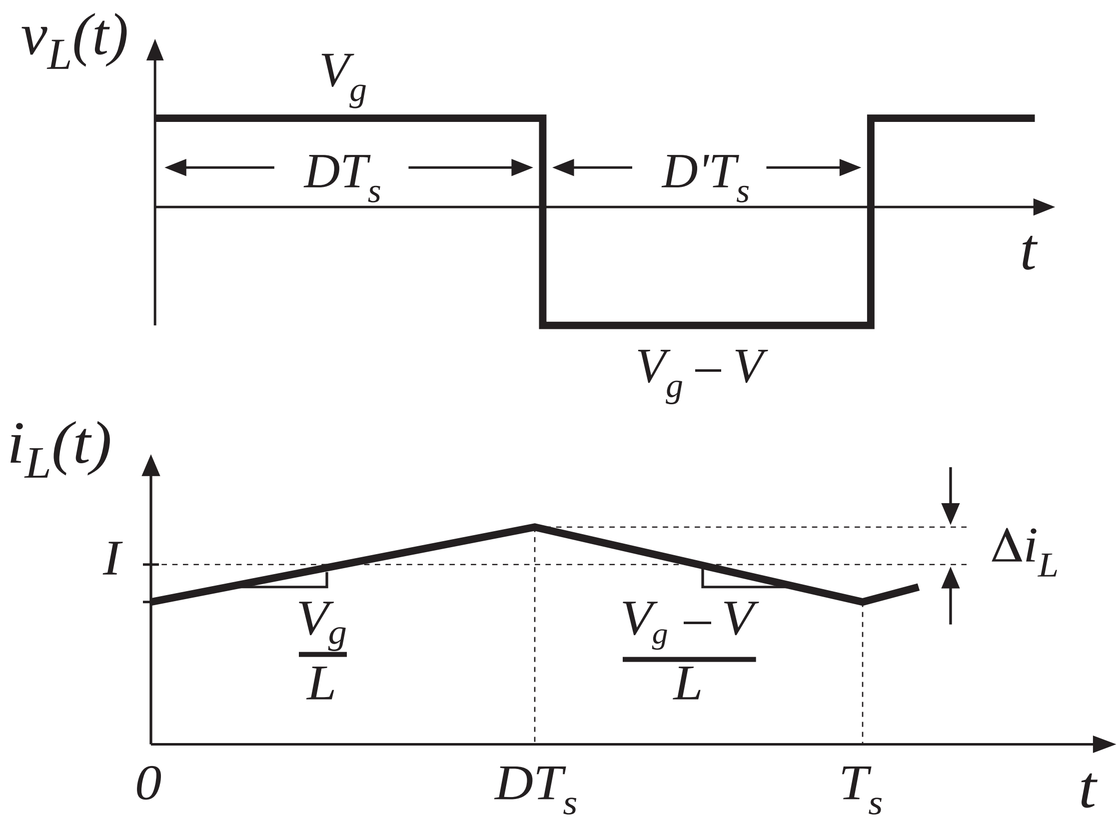
$$v_L(t) = L\,\frac{di_L(t)}{dt}$$

Solve for the slope:

$$\frac{di_L(t)}{dt} = \frac{v_L(t)}{L} \approx \frac{V_g - V}{L}$$

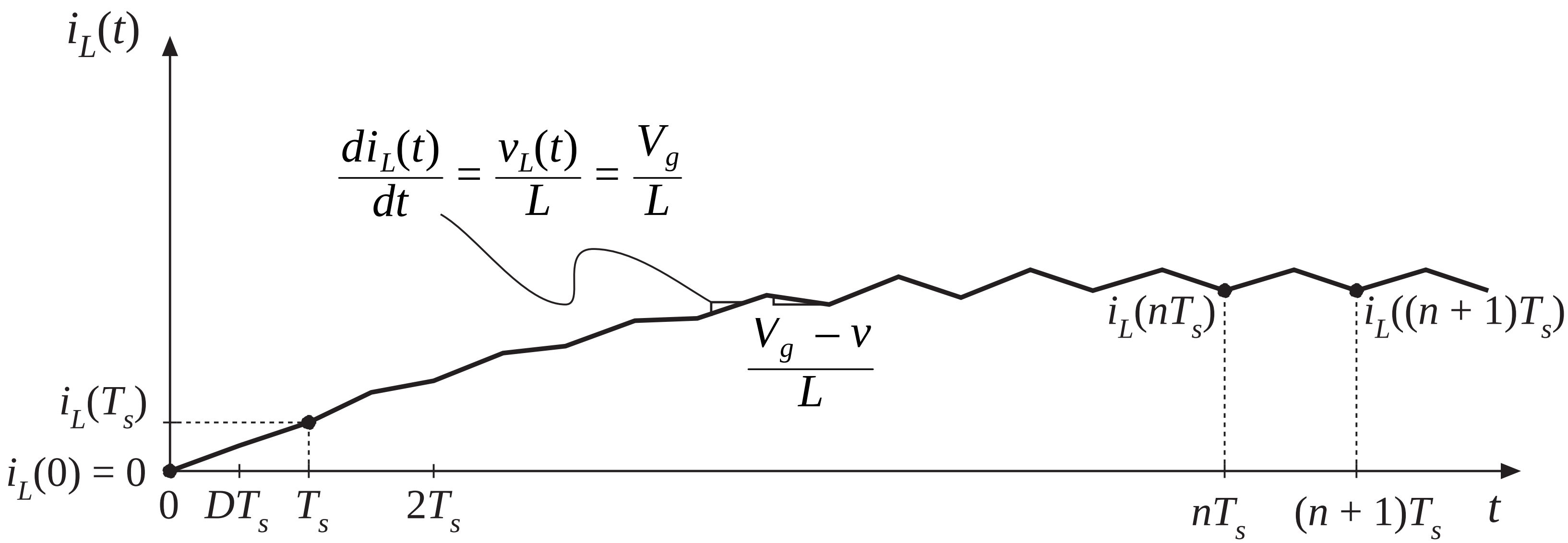
$\Rightarrow$ *The inductor current changes with an essentially constant, but different, slope*

Inductor voltage and current waveforms



$$\frac{di_L(t)}{dt} = \frac{v_L(t)}{L}$$

Inductor current waveform during turn-on transient



When the converter operates in equilibrium:

$$i_L((n + 1)T_s) = i_L(nT_s)$$

The principle of inductor volt-second balance: Derivation

Inductor defining relation:

$$v_L(t) = L \frac{di_L(t)}{dt}$$

Integrate over one complete switching period:

$$i_L(T_s) - i_L(0) = \frac{1}{L} \int_0^{T_s} v_L(t)\, dt$$

In periodic steady state, the net change in inductor current is zero:

$$0 = \int_0^{T_s} v_L(t)\, dt$$

Hence, the total area (or volt-seconds) under the inductor voltage waveform is zero whenever the converter operates in steady state.

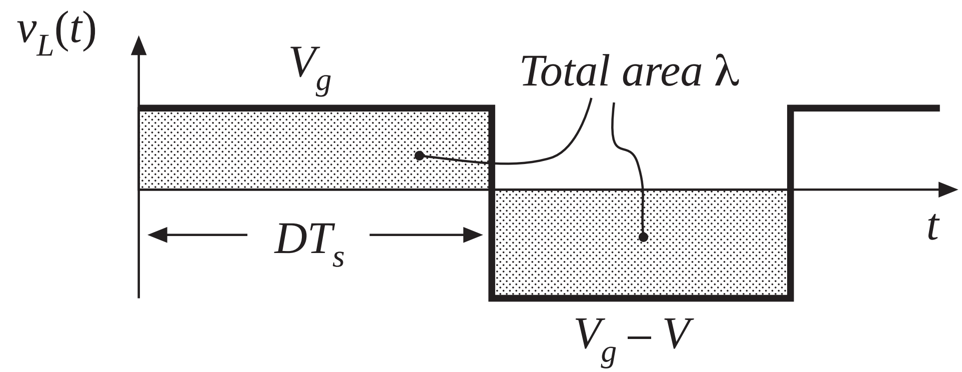
An equivalent form:

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

The average inductor voltage is zero in steady state.

Inductor volt-second balance: Boost converter example

Inductor voltage waveform, previously derived:



Integral of voltage waveform is area of rectangles:

$$\lambda = \int_0^{T_s} v_L(t)dt = \left(V_g\right)\left(DT_s\right) + \left(V_g - V\right)\left(D'T_s\right) \qquad \text{with } D' = 1 - D$$

Average voltage is

$$\left\langle v_L \right\rangle = \frac{\lambda}{T_s} = D\left(V_g\right) + D'\left(V_g - V\right)$$

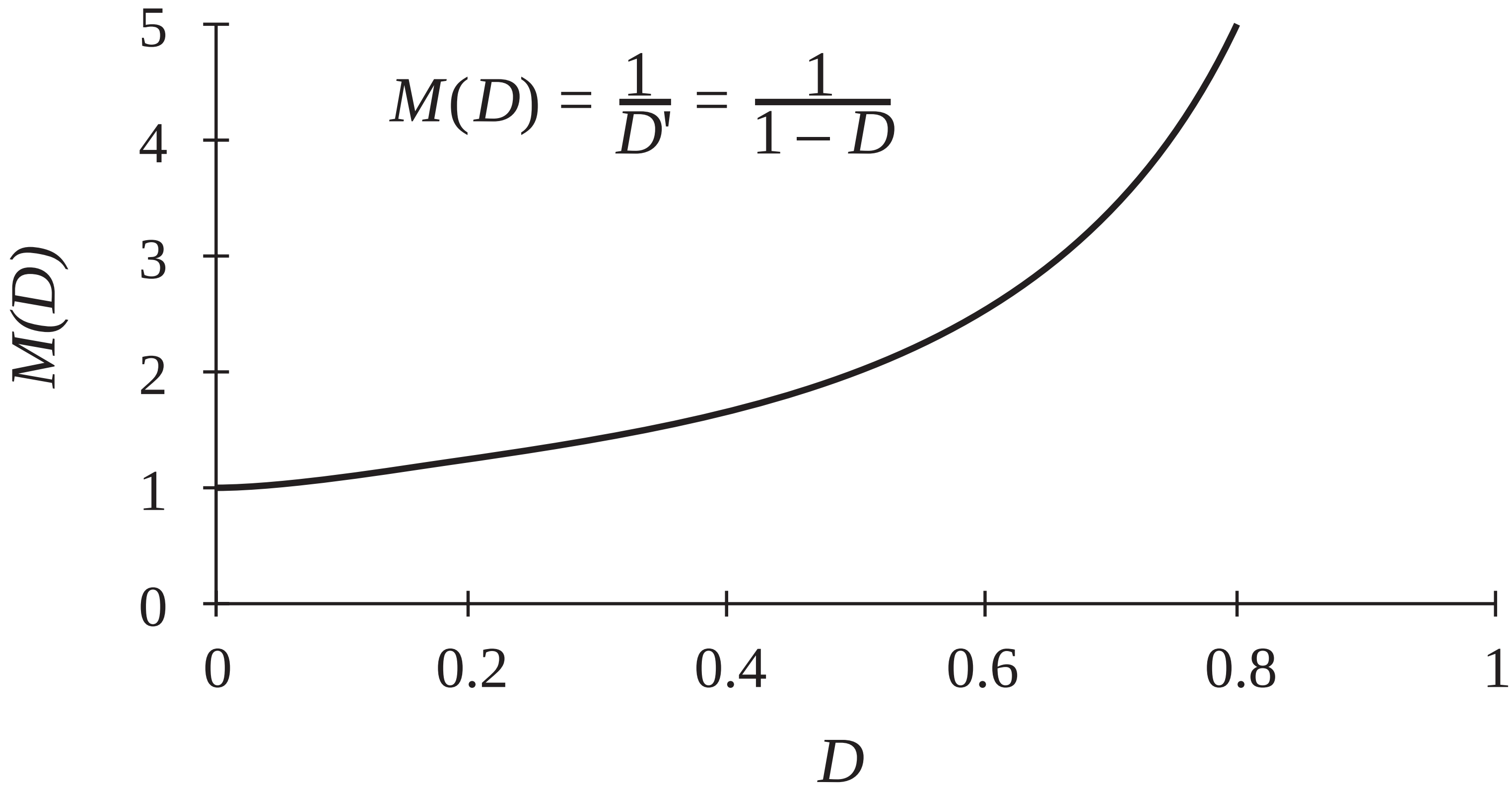
Equate to zero and solve for *V:*

$$V_g\,(D + D') - V\,D' = 0 \quad \Rightarrow \quad V = \frac{V_g}{D'}$$

The voltage conversion ratio is

$$M(D) = \frac{V}{V_g} = \frac{1}{D'} = \frac{1}{1 - D}$$

Conversion ratio *M(D)* of the boost converter



The principle of capacitor charge balance: Derivation

Capacitor defining relation:

$$i_C(t) = C \frac{dv_C(t)}{dt}$$

Integrate over one complete switching period:

$$v_C(T_s) - v_C(0) = \frac{1}{C}\int_0^{T_s} i_C(t)\, dt$$

In periodic steady state, the net change in capacitor voltage is zero:

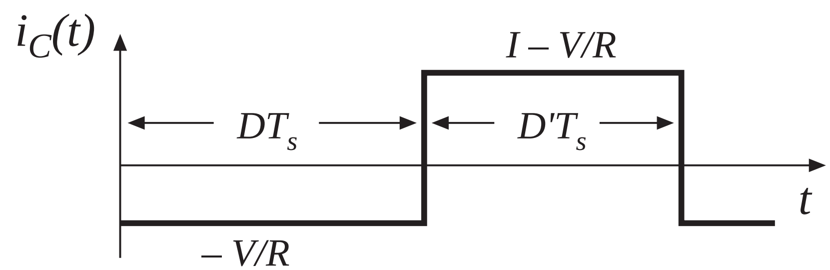
$$0 = \frac{1}{T_s}\int_0^{T_s} i_C(t)\, dt = \left\langle i_C \right\rangle$$

Hence, the total area (or charge) under the capacitor current waveform is zero whenever the converter operates in steady state. The average capacitor current is then zero.

Determination of inductor current dc component
using capacitor charge balance

Capacitor charge balance:

$$\int_0^{T_s} i_C(t)\,dt = \left(-\frac{V}{R}\right) DT_s + \left(I - \frac{V}{R}\right) D'T_s$$



Collect terms and equate to zero:

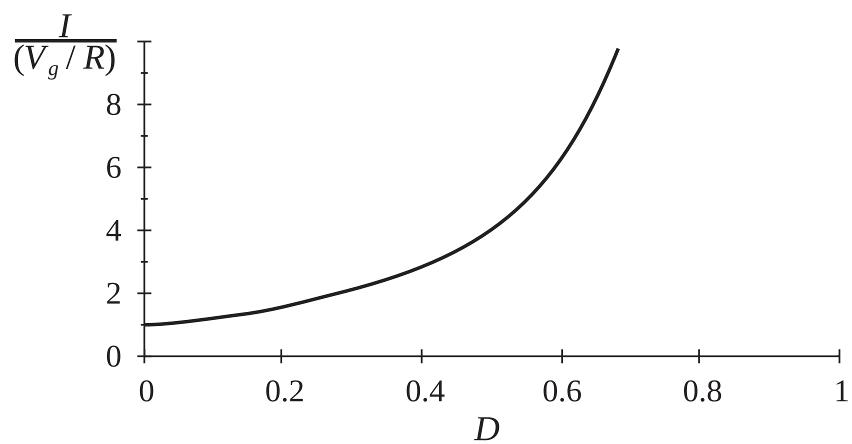
$$-\frac{V}{R}(D + D') + I\,D' = 0$$

Solve for I:

$$I = \frac{V}{D'\,R}$$

Eliminate V to express in terms of V_g:

$$I = \frac{V_g}{D'^2\,R}$$



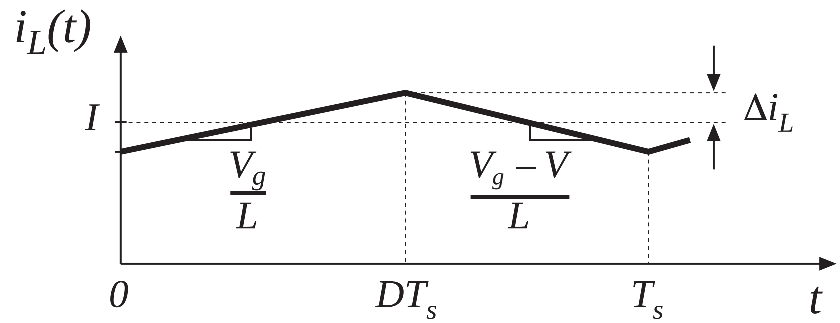
Determination of inductor current ripple

Inductor current slope during subinterval 1:

$$\frac{di_L(t)}{dt} = \frac{v_L(t)}{L} = \frac{V_g}{L}$$

Inductor current slope during subinterval 2:

$$\frac{di_L(t)}{dt} = \frac{v_L(t)}{L} = \frac{V_g - V}{L}$$



Change in inductor current during subinterval 1 is (*slope*) (*length of subinterval*):

$$2\Delta i_L = \frac{V_g}{L} DT_s$$

Solve for peak ripple:

$$\Delta i_L = \frac{V_g}{2L} DT_s$$

- Choose *L* such that desired ripple magnitude is obtained

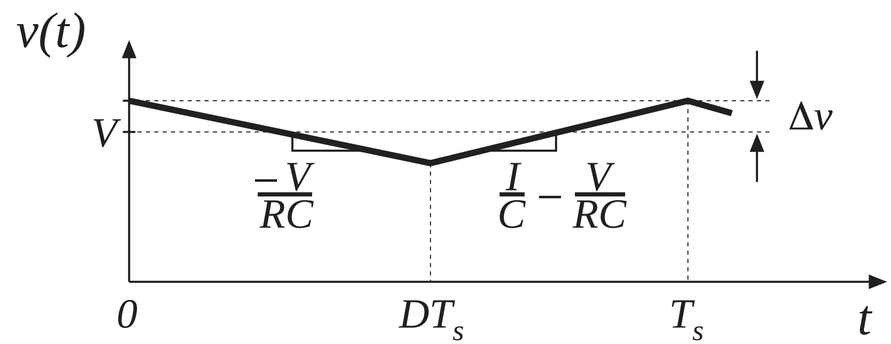
Determination of capacitor voltage ripple

Capacitor voltage slope during subinterval 1:

$$\frac{dv_C(t)}{dt} = \frac{i_C(t)}{C} = \frac{-V}{RC}$$

Capacitor voltage slope during subinterval 2:

$$\frac{dv_C(t)}{dt} = \frac{i_C(t)}{C} = \frac{I}{C} - \frac{V}{RC}$$



Change in capacitor voltage during subinterval 1 is (*slope*) (*length of subinterval*):

$$-2\Delta v = \frac{-V}{RC} DT_s$$

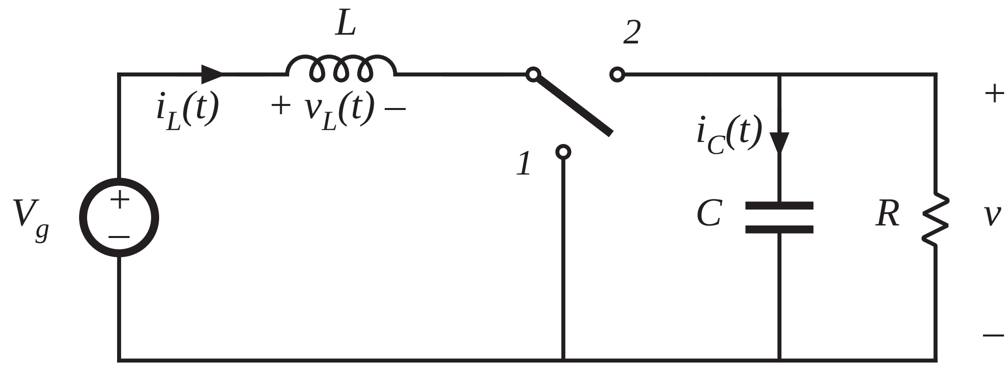
Solve for peak ripple:

$$\Delta v = \frac{V}{2RC} DT_s$$

- Choose C such that desired voltage ripple magnitude is obtained
- In practice, capacitor *equivalent series resistance* (esr) leads to increased voltage ripple

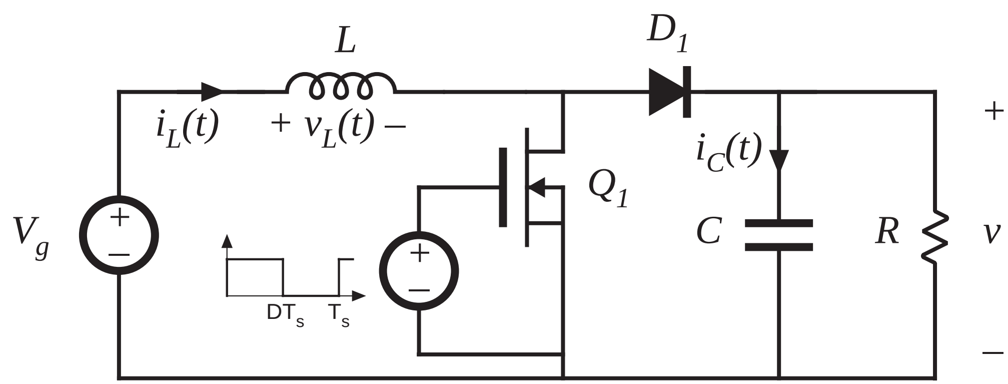
Realization of SPDT switch using transistor and diode

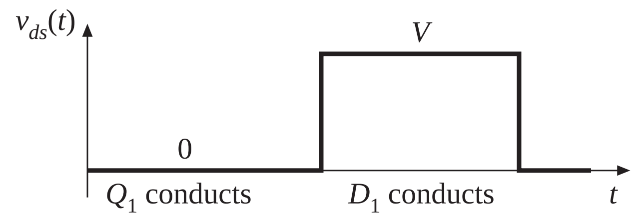
Boost converter with ideal SPDT switch



Realization using power MOSFET and diode

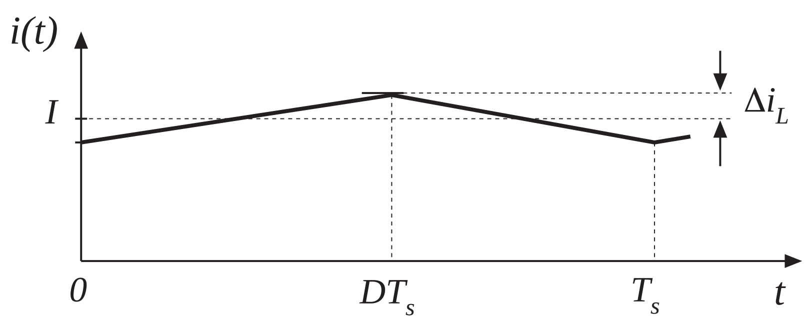
Inductor current forward-biases diode when MOSFET is off

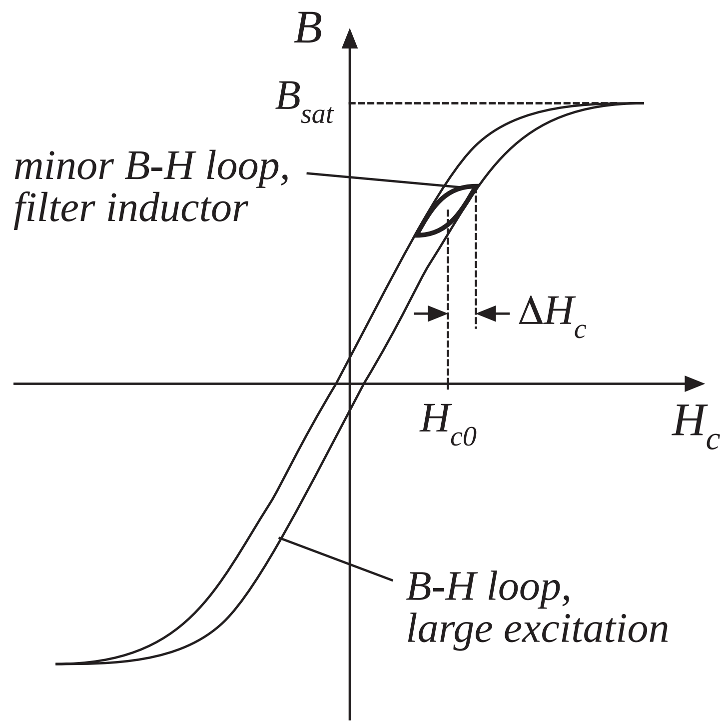




Designing the filter inductor

Boost example: inductor current waveform

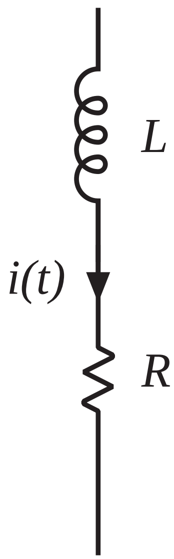




In an inductor, magnetic field $H(t)$ is proportional to winding current $i(t)$ via Ampere's law. Flux density $B(t)$ is proportional to integral of winding voltage $v(t)$ through Faraday's law.

Must avoid saturation of core: $B(t) < B_{sat}$.

Filter inductor design constraints



Objective:

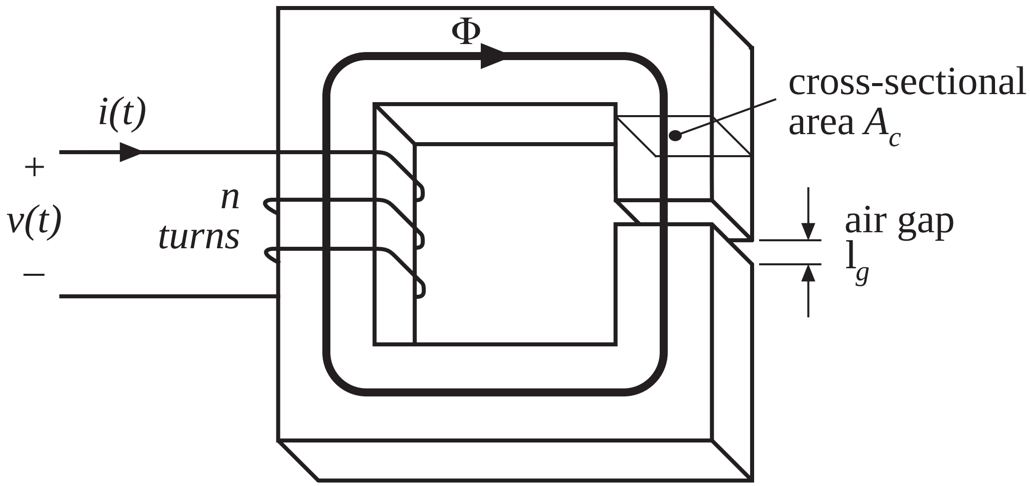
Design inductor having a given inductance L,

which carries worst-case current I_{max} without saturating,

and which has a given winding resistance R, or, equivalently, exhibits a worst-case copper loss of

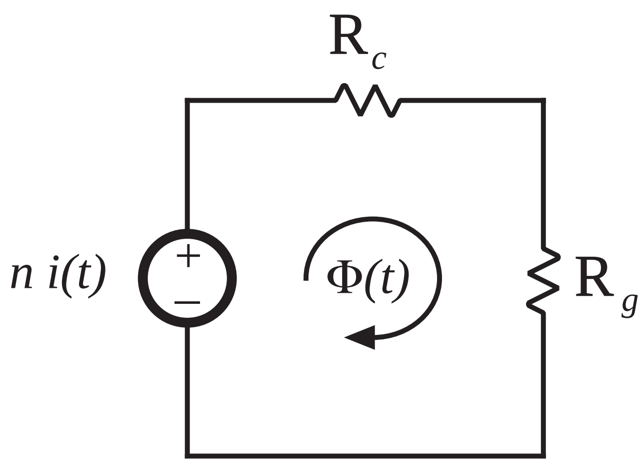
$$P_{cu} = I_{rms}^{2} R$$

Assumed filter inductor geometry



$$\mathcal{R}_c = \frac{l_c}{\mu_c A_c}$$

$$\mathcal{R}_g = \frac{l_g}{\mu_0 A_c}$$



Solve magnetic circuit:

$$ni = \Phi\left(\mathcal{R}_c + \mathcal{R}_g\right)$$

For $\mathcal{R}_c > \mathcal{R}_g$: $\quad ni \approx \Phi\, \mathcal{R}_g$

First constraint: maximum flux density

Given a peak winding current I_{max}, it is desired to operate the core flux density at a peak value B_{max}. The value of B_{max} is chosen to be less than the worst-case saturation flux density of the core material, B_{sat}.

From solution of magnetic circuit:

$$ni = BA_c \mathcal{R}_g$$

Let $I = I_{max}$ and $B = B_{max}$:

$$nI_{max} = B_{max} A_c \mathcal{R}_g = B_{max} \frac{l_g}{\mu_0}$$

This is constraint #1. The turns ratio n and air gap length l_g are unknown.

Second constraint: obtain desired inductance

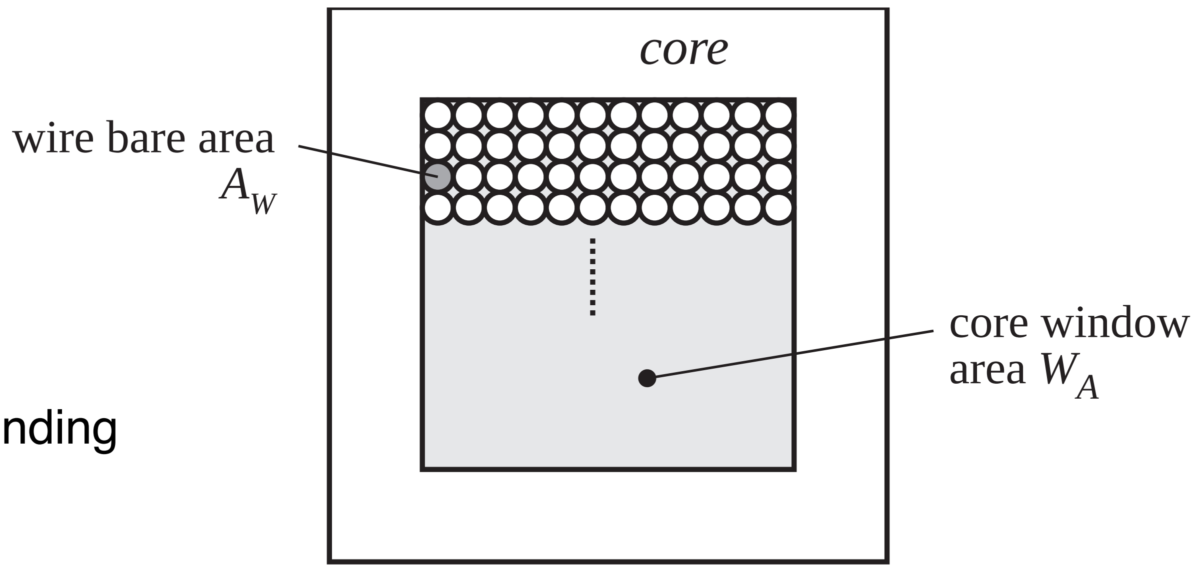
Must obtain specified inductance L. We know that the inductance is

$$L = \frac{n^2}{\mathcal{R}_g} = \frac{\mu_0 A_c n^2}{\ell_g}$$

This is constraint #2. The turns ratio n, core area A_c, and air gap length ℓ_g are unknown.

Third constraint: winding area

Wire must fit through core window (i.e., hole in center of core)



Total area of copper in window:

nA_W

Area available for winding conductors:

$K_u W_A$

Third design constraint:

$K_u W_A \geq nA_W$

The window utilization factor K_u

also called the "fill factor"

K_u is the fraction of the core window area that is filled by copper

Mechanisms that cause K_u to be less than 1:

- Round wire does not pack perfectly, which reduces K_u by a factor of 0.7 to 0.55 depending on winding technique
- Insulation reduces K_u by a factor of 0.95 to 0.65, depending on wire size and type of insulation
- Bobbin uses some window area
- Additional insulation may be required between windings

Typical values of K_u :

0.5 for simple low-voltage inductor

0.25 to 0.3 for off-line transformer

0.05 to 0.2 for high-voltage transformer (multiple kV)

0.65 for low-voltage foil-winding inductor

Fourth constraint: winding resistance

The resistance of the winding is

$$R = \rho \, \frac{\ell_b}{A_W}$$

where ρ is the resistivity of the conductor material, ℓ_b is the length of the wire, and A_W is the wire bare area. The resistivity of copper at room temperature is $1.724 \cdot 10^{-6}\ \Omega\text{-cm}$. The length of the wire comprising an n-turn winding can be expressed as

$$\ell_b = n\,(MLT)$$

where (MLT) is the mean-length-per-turn of the winding. The mean-length-per-turn is a function of the core geometry. The above equations can be combined to obtain the fourth constraint:

$$R = \rho \, \frac{n\,(MLT)}{A_W}$$

Combine all four constraints: The core geometrical constant K_g

The four constraints:

$$nI_{max} = B_{max} \frac{\ell_g}{\mu_0}$$

$$L = \frac{\mu_0 A_c n^2}{\ell_g}$$

$$K_u W_A \geq n A_W$$

$$R = \rho \frac{n\,(MLT)}{A_W}$$

These equations involve the quantities

A_c, W_A, and MLT, which are functions of the core geometry,

I_{max}, B_{max} , μ_0, L, K_u, R, and ρ, which are given specifications or other known quantities, and

n, ℓ_g, and A_W, which are unknowns.

Eliminate the three unknowns, leading to a single equation involving the remaining quantities.

Core geometrical constant K_g

Elimination of n, ℓ_g, and A_W leads to

$$\frac{A_c^2 W_A}{(MLT)} \geq \frac{\rho L^2 I_{max}^2}{B_{max}^2 R K_u}$$

- Right-hand side: specifications or other known quantities
- Left-hand side: function of only core geometry

So we must choose a core whose geometry satisfies the above equation.

The core geometrical constant K_g is defined as

$$K_g = \frac{A_c^2 W_A}{(MLT)}$$

Discussion

$$K_g = \frac{A_c^2 W_A}{(MLT)} \geq \frac{\rho L^2 I_{max}^2}{B_{max}^2 R K_u}$$

K_g is a figure-of-merit that describes the effective electrical size of magnetic cores, in applications where the following quantities are specified:

- Copper loss
- Maximum flux density

How specifications affect the core size:

A smaller core can be used by increasing

$B_{max} \Rightarrow$ use core material having higher B_{sat}

$R \Rightarrow$ allow more copper loss

How the core geometry affects electrical capabilities:

A larger K_g can be obtained by increase of

$A_c \Rightarrow$ more iron core material, or

$W_A \Rightarrow$ larger window and more copper

A step-by-step procedure

The following quantities are specified, using the units noted:

Wire resistivity	ρ	(Ω-cm)
Peak winding current	I_{max}	(A)
Inductance	L	(H)
Winding resistance	R	(Ω)
Winding fill factor	K_u	
Core maximum flux density	B_{max}	(T)

The core dimensions are expressed in cm:

Core cross-sectional area	A_c	(cm^2)
Core window area	W_A	(cm^2)
Mean length per turn	MLT	(cm)

The use of centimeters rather than meters requires that appropriate factors be added to the design equations.

Determine core size

$$K_g \geq \frac{\rho L^2 I_{max}^2}{B_{max}^2 R K_u} 10^8 \qquad (\mathrm{cm}^5)$$

Choose a core which is large enough to satisfy this inequality (*see magnetics design tables for lists of K_g of standard core geometries*).

Note the values of A_c, W_A, and MLT for this core.

(In this lab, we will use the core supplied in the parts kits, which is a PQ32/20 core. The data for this core size is listed in the magnetics design tables)

Determine air gap length

$$\ell_g = \frac{\mu_0 L I_{max}^2}{B_{max}^2 A_c} 10^4 \qquad \text{(m)}$$

with A_c expressed in cm^2. $\mu_0 = 4\pi 10^{-7}$ H/m.

The air gap length is given in meters.

The value expressed above is approximate, and neglects fringing flux and other nonidealities.

Determine number of turns n

$$n = \frac{LI_{max}}{B_{max}A_c}\ 10^4$$

Evaluate wire size

$$A_W \le \frac{K_u W_A}{n} \qquad (\text{cm}^2)$$

Select wire with bare copper area A_W less than or equal to this value. An American Wire Gauge table is included on the course web site.

As a check, the winding resistance can be computed:

$$R = \frac{\rho n\,(MLT)}{A_w} \qquad (\Omega)$$

The winding power loss must lead to an acceptable temperature rise. It may take many minutes, or even hours, for the core to reach thermal equilibrium. Thermal runaway can occur if the winding resistance is too large.

See core tables for thermal resistances of standard core geometries. For the PQ32/20, R_{th} is approximately 15 ˚C/Watt.

Appendix 2

Magnetics Design Tables

Geometrical data for several standard ferrite core shapes are listed here. The geometrical constant K_g is a measure of core size, useful for designing inductors and transformers which attain a given copper loss [1]. The K_g method for inductor design is described in Chapter 13. K_g is defined as

$$K_g = \frac{A_c^2 W_A}{MLT} \tag{A2.1}$$

where A_c is the core cross-sectional area, W_A is the window area, and MLT is the winding mean-length-per-turn. The geometrical constant K_{gfe} is a similar measure of core size, which is useful for designing ac inductors and transformers when the total copper plus core loss is constrained. The K_{gfe} method for magnetics design is described in Chapter 14. K_{gfe} is defined as

$$K_{gfe} = \frac{W_A A_c^{2(1-1/\beta)}}{MLT\ l_e^{2/\beta}} u(\beta) \tag{A2.2}$$

where l_e is the core mean magnetic path length, and β is the core loss exponent:

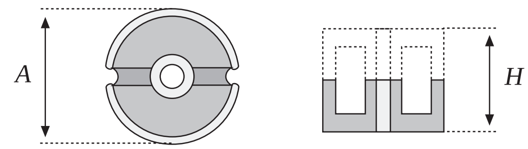
$$P_{fe} = K_{fe} B_{\max}^{\beta} \tag{A2.3}$$

For modern ferrite materials, β typically lies in the range 2.6 to 2.8. The quantity $u(\beta)$ is defined as

$$u(\beta) = \left[\left(\frac{\beta}{2}\right)^{-\left(\frac{\beta}{\beta+2}\right)} + \left(\frac{\beta}{2}\right)^{\left(\frac{2}{\beta+2}\right)} \right]^{-\left(\frac{\beta+2}{\beta}\right)} \tag{A2.4}$$

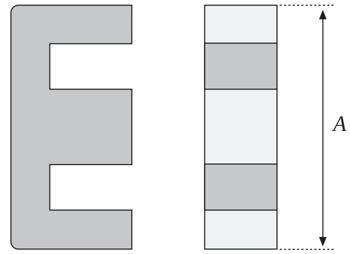
$u(\beta)$ is equal to 0.305 for $\beta = 2.7$. This quantity varies by roughly 5% over the range $2.6 \le \beta \le 2.8$. Values of K_{gfe} are tabulated for $\beta = 2.7$; variation of K_{gfe} over the range $2.6 \le \beta \le 2.8$ is typically quite small.

A2.1 Pot core data



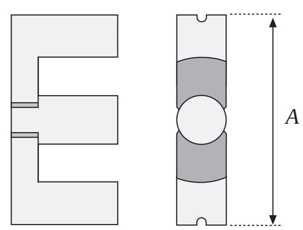
Core type (AH) (mm)	Geometrical constant K_g cm^5	Geometrical constant K_{gfe} cm^x	Cross-sectional area A_c (cm^2)	Bobbin winding area W_A (cm^2)	Mean length per turn MLT (cm)	Magnetic path length l_m (cm)	Thermal resistance R_{th} (˚C/W)	Core weight (g)
704	$0.738 \cdot 10^{-6}$	$1.61 \cdot 10^{-6}$	0.070	$0.22 \cdot 10^{-3}$	1.46	1.0		0.5
905	$0.183 \cdot 10^{-3}$	$256 \cdot 10^{-6}$	0.101	0.034	1.90	1.26		1.0
1107	$0.667 \cdot 10^{-3}$	$554 \cdot 10^{-6}$	0.167	0.055	2.30	1.55		1.8
1408	$2.107 \cdot 10^{-3}$	$1.1 \cdot 10^{-3}$	0.251	0.097	2.90	2.00	100	3.2
1811	$9.45 \cdot 10^{-3}$	$2.6 \cdot 10^{-3}$	0.433	0.187	3.71	2.60	60	7.3
2213	$27.1 \cdot 10^{-3}$	$4.9 \cdot 10^{-3}$	0.635	0.297	4.42	3.15	38	13
2616	$69.1 \cdot 10^{-3}$	$8.2 \cdot 10^{-3}$	0.948	0.406	5.28	3.75	30	20
3019	0.180	$14.2 \cdot 10^{-3}$	1.38	0.587	6.20	4.50	23	34
3622	0.411	$21.7 \cdot 10^{-3}$	2.02	0.748	7.42	5.30	19	57
4229	1.15	$41.1 \cdot 10^{-3}$	2.66	1.40	8.60	6.81	13.5	104

A2.2 EE core data



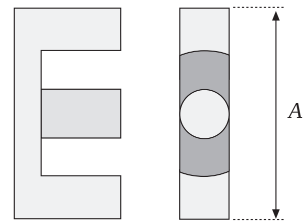
Core type (A) (mm)	Geometrical constant K_g cm^5	Geometrical constant K_{gfe} cm^x	Cross-sectional area A_c (cm^2)	Bobbin winding area W_A (cm^2)	Mean length per turn MLT (cm)	Magnetic path length l_m (cm)	Core weight (g)
EE12	$0.731 \cdot 10^{-3}$	$0.458 \cdot 10^{-3}$	0.14	0.085	2.28	2.7	2.34
EE16	$2.02 \cdot 10^{-3}$	$0.842 \cdot 10^{-3}$	0.19	0.190	3.40	3.45	3.29
EE19	$4.07 \cdot 10^{-3}$	$1.3 \cdot 10^{-3}$	0.23	0.284	3.69	3.94	4.83
EE22	$8.26 \cdot 10^{-3}$	$1.8 \cdot 10^{-3}$	0.41	0.196	3.99	3.96	8.81
EE30	$85.7 \cdot 10^{-3}$	$6.7 \cdot 10^{-3}$	1.09	0.476	6.60	5.77	32.4
EE40	0.209	$11.8 \cdot 10^{-3}$	1.27	1.10	8.50	7.70	50.3
EE50	0.909	$28.4 \cdot 10^{-3}$	2.26	1.78	10.0	9.58	116
EE60	1.38	$36.4 \cdot 10^{-3}$	2.47	2.89	12.8	11.0	135
EE70/68/19	5.06	$127 \cdot 10^{-3}$	3.24	6.75	14.0	9.0	280

A2.3 EC core data



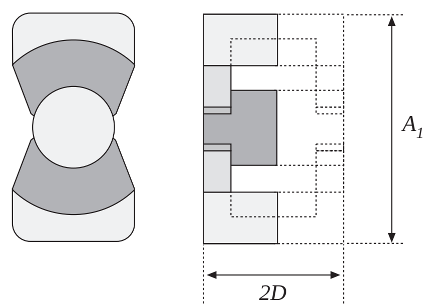
Core type (A) (mm)	Geometrical constant K_g cm^5	Geometrical constant K_{gfe} cm^x	Cross-sectional area A_c (cm^2)	Bobbin winding area W_A (cm^2)	Mean length per turn MLT (cm)	Magnetic path length l_m (cm)	Thermal resistance R_{th} (°C/W)	Core weight (g)
EC35	0.131	$9.9 \cdot 10^{-3}$	0.843	0.975	5.30	7.74	18.5	35.5
EC41	0.374	$19.5 \cdot 10^{-3}$	1.21	1.35	5.30	8.93	16.5	57.0
EC52	0.914	$31.7 \cdot 10^{-3}$	1.80	2.12	7.50	10.5	11.0	111
EC70	2.84	$56.2 \cdot 10^{-3}$	2.79	4.71	12.9	14.4	7.5	256

A2.4 ETD core data



Core type (A) (mm)	Geometrical constant K_g cm^5	Geometrical constant K_{gfe} cm^x	Cross-sectional area A_c (cm^2)	Bobbin winding area W_A (cm^2)	Mean length per turn MLT (cm)	Magnetic path length l_m (cm)	Thermal resistance R_{th} (°C/W)	Core weight (g)
ETD29	0.0978	$8.5 \cdot 10^{-3}$	0.76	0.903	5.33	7.20		30
ETD34	0.193	$13.1 \cdot 10^{-3}$	0.97	1.23	6.00	7.86	19	40
ETD39	0.397	$19.8 \cdot 10^{-3}$	1.25	1.74	6.86	9.21	15	60
ETD44	0.846	$30.4 \cdot 10^{-3}$	1.74	2.13	7.62	10.3	12	94
ETD49	1.42	$41.0 \cdot 10^{-3}$	2.11	2.71	8.51	11.4	11	124

A2.5 PQ core data



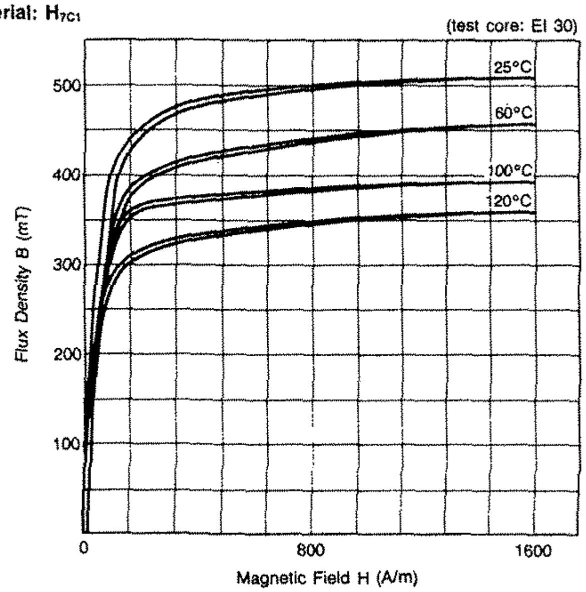
Core type ($A_1/2D$) (mm)	Geometrical constant K_g cm^5	Geometrical constant K_{gfe} cm^x	Cross-sectional area A_c (cm^2)	Bobbin winding area W_A (cm^2)	Mean length per turn MLT (cm)	Magnetic path length l_m (cm)	Core weight (g)
PQ 20/16	$22.4 \cdot 10^{-3}$	$3.7 \cdot 10^{-3}$	0.62	0.256	4.4	3.74	13
PQ 20/20	$33.6 \cdot 10^{-3}$	$4.8 \cdot 10^{-3}$	0.62	0.384	4.4	4.54	15
PQ 26/20	$83.9 \cdot 10^{-3}$	$7.2 \cdot 10^{-3}$	1.19	0.333	5.62	4.63	31
PQ 26/25	0.125	$9.4 \cdot 10^{-3}$	1.18	0.503	5.62	5.55	36
PQ 32/20	0.203	$11.7 \cdot 10^{-3}$	1.70	0.471	6.71	5.55	42
PQ 32/30	0.384	$18.6 \cdot 10^{-3}$	1.61	0.995	6.71	7.46	55
PQ 35/35	0.820	$30.4 \cdot 10^{-3}$	1.96	1.61	7.52	8.79	73
PQ 40/40	1.20	$39.1 \cdot 10^{-3}$	2.01	2.50	8.39	10.2	95

A2.6 American wire gauge data

AWG#	Bare area, 10^{-3} cm^2	Resistance, 10^{-6} Ω/cm	Diameter, cm
0000	1072.3	1.608	1.168
000	850.3	2.027	1.040
00	674.2	2.557	0.927
0	534.8	3.224	0.825
1	424.1	4.065	0.735
2	336.3	5.128	0.654
3	266.7	6.463	0.583
4	211.5	8.153	0.519
5	167.7	10.28	0.462
6	133.0	13.0	0.411
7	105.5	16.3	0.366
8	83.67	20.6	0.326
9	66.32	26.0	0.291
10	52.41	32.9	0.267
11	41.60	41.37	0.238
12	33.08	52.09	0.213
13	26.26	69.64	0.190
14	20.02	82.80	0.171
15	16.51	104.3	0.153
16	13.07	131.8	0.137
17	10.39	165.8	0.122
18	8.228	209.5	0.109
19	6.531	263.9	0.0948
20	5.188	332.3	0.0874
21	4.116	418.9	0.0785
22	3.243	531.4	0.0701
23	2.508	666.0	0.0632
24	2.047	842.1	0.0566
25	1.623	1062.0	0.0505
26	1.280	1345.0	0.0452
27	1.021	1687.6	0.0409
28	0.8046	2142.7	0.0366
29	0.6470	2664.3	0.0330
30	0.5067	3402.2	0.0294
31	0.4013	4294.6	0.0267
32	0.3242	5314.9	0.0241
33	0.2554	6748.6	0.0236
34	0.2011	8572.8	0.0191
35	0.1589	10849	0.0170
36	0.1266	13608	0.0152
37	0.1026	16801	0.0140
38	0.08107	21266	0.0124
39	0.06207	27775	0.0109
40	0.04869	35400	0.0096
41	0.03972	43405	0.00863
42	0.03166	54429	0.00762
43	0.02452	70308	0.00685

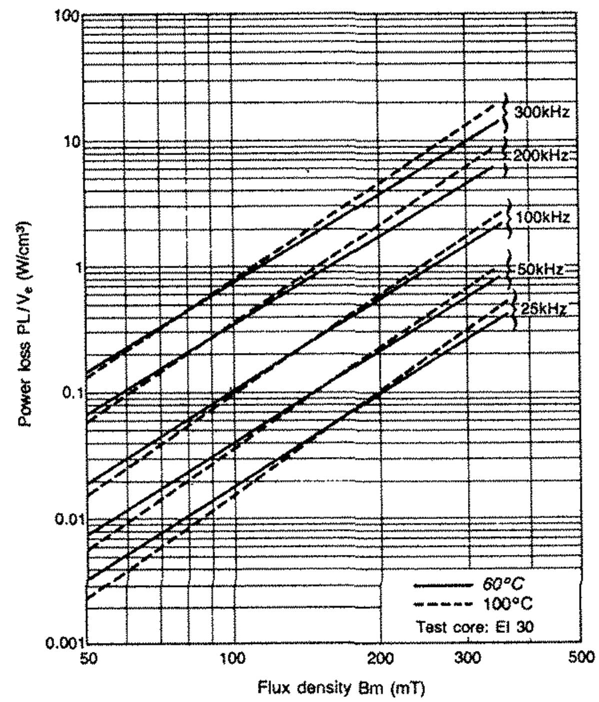
Magnetization Curves (Typical)

Material: H_{7C1}



Power loss (Typical)

Material: H_{7C1}



Boost DC-DC Switching Power Converter

Pre-lab assignment #1

ECEN 4517 / 5017

1. Reading material:

- Introduction to converters handout
- Inductor design handout
- Laboratory procedure
- Supplementary material: Tutorial article on dc-dc converters; Two sample converter problems with solutions; Slides on basic magnetics; Data Sheets for components used on the converter test board.

2. In the laboratory, you are going to construct and test a boost DC-DC power converter that operates at the following nominal quiescent point: input DC voltage V_g=50V, output DC voltage V = 100V, output load P_{load}=100W, switching frequency f_s=100kHz. The converter circuit is shown in the figure below.

(a) Choose the inductance L such that the peak current ripple Δi_L is 10% of the dc inductor current I_L. You can neglect all losses in the converter.

(b) For the inductance L selected in (a), find, sketch and label the steady-state waveforms $i_L(t)$, $i_t(t)$, $i_d(t)$, and $v_t(t)$ during one switching period. Find the peak ripple Δv in the output voltage. You can neglect all losses in the converter.

(c) Design the inductor using PQ32/20 core with the core cross-sectional area A_c=1.7cm^2, the bobbin winding area W_A=0.471cm^2, and mean length per turn MLT=6.71cm. Use a peak flux density B_{max}=0.2T, and assume a fill factor K_u=0.5. Select: number of turns n, wire gauge, and air gap length l_g. Find the winding resistance R_l and the copper loss P_{cu} on the inductor.

The due date for the prelab assignment is posted on the course Web page. Keep a copy of your work for use during the lab experiment.