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DIPARTIMENTO DI ELETTRONICA,
INFORMAZIONE E BIOINGEGNERIA





Power Electronics

Part 8

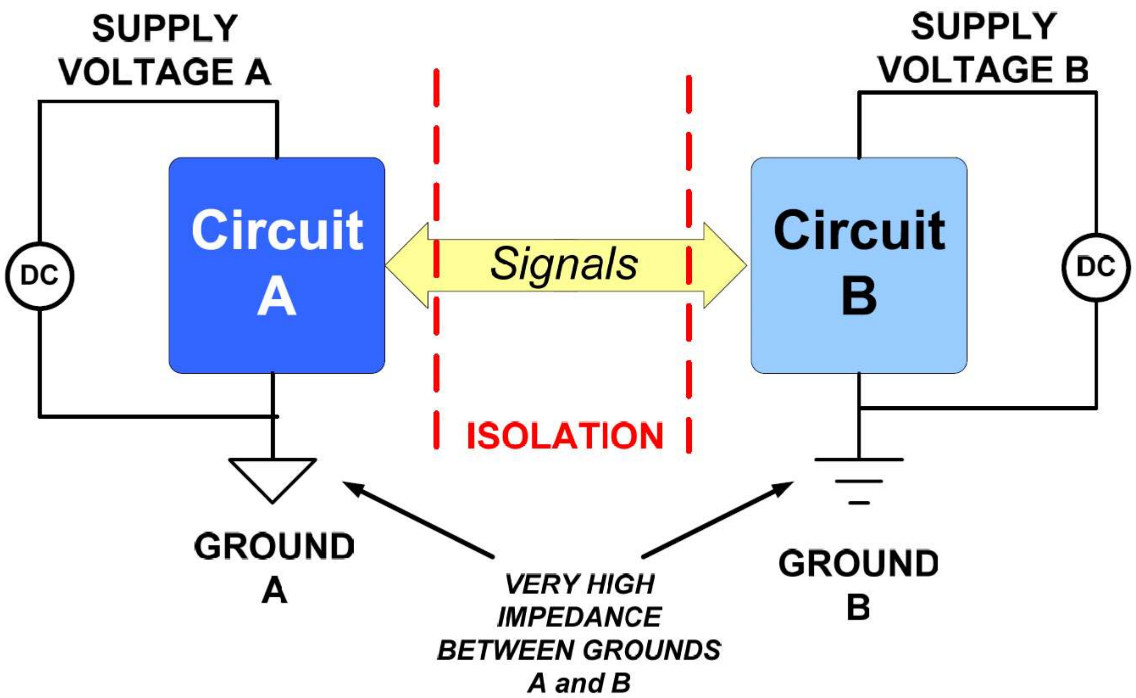
Massimo Ghioni

Isolated DC-DC converters

What is isolation?

Galvanic isolation

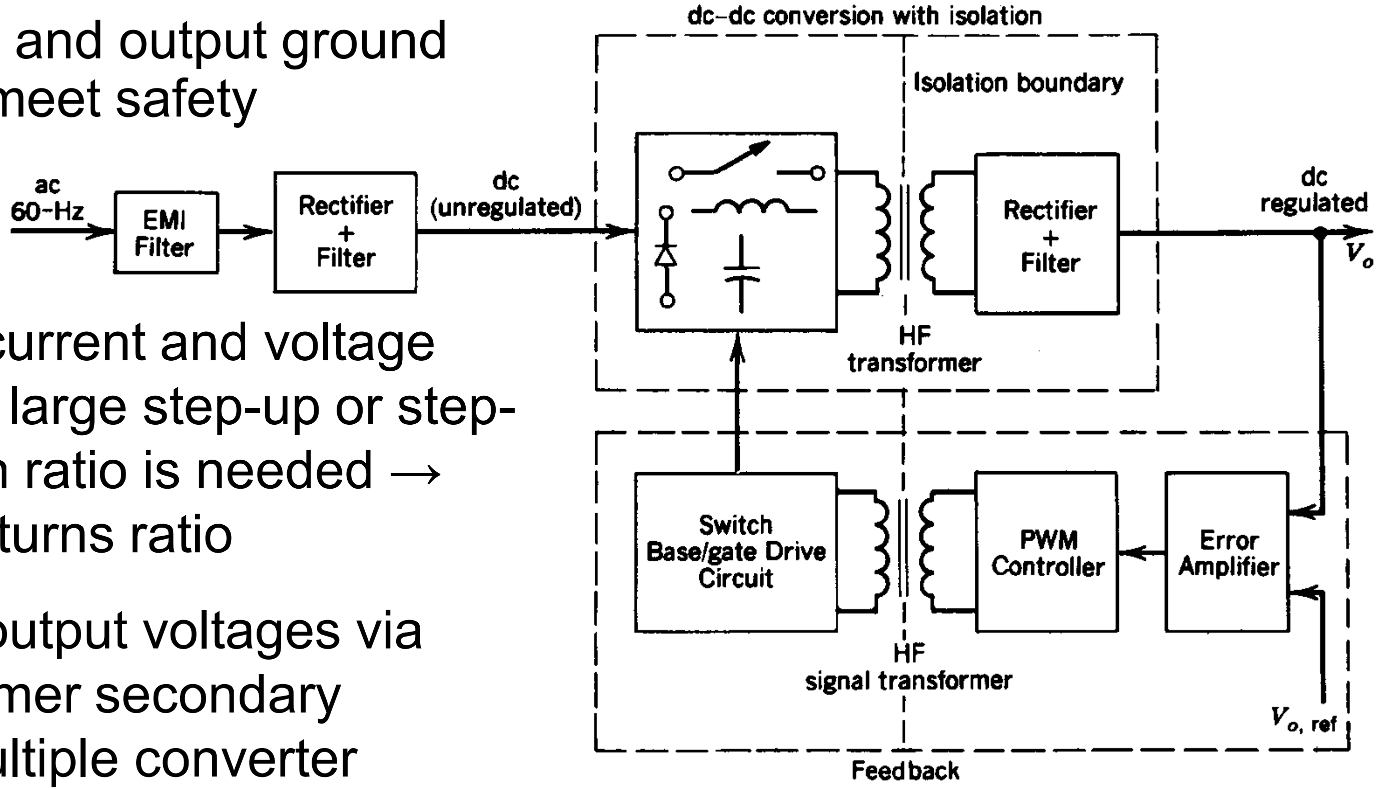
- No current flow from one circuit to another
- Signal information passes from one circuit to another



Isolated DC-DC converters

Objectives

- Isolation of input and output ground connections, to meet safety requirements



- Minimization of current and voltage stresses when a large step-up or step-down conversion ratio is needed → use transformer turns ratio
- Obtain multiple output voltages via multiple transformer secondary windings and multiple converter secondary circuits
- Reduction of transformer size by incorporating high frequency isolation transformer inside converter

Note: this is only one possible arrangement. Controller often on primary side with switch.

Multiple outputs

- Transformer allows multiple outputs to be “quasi-regulated” using a single controller and feedback loop.

- e.g. most PC power supplies achieve multiple outputs this way.

- In practice, feedback loop uses a “weighted average” of output voltages to achieve best overall regulation.

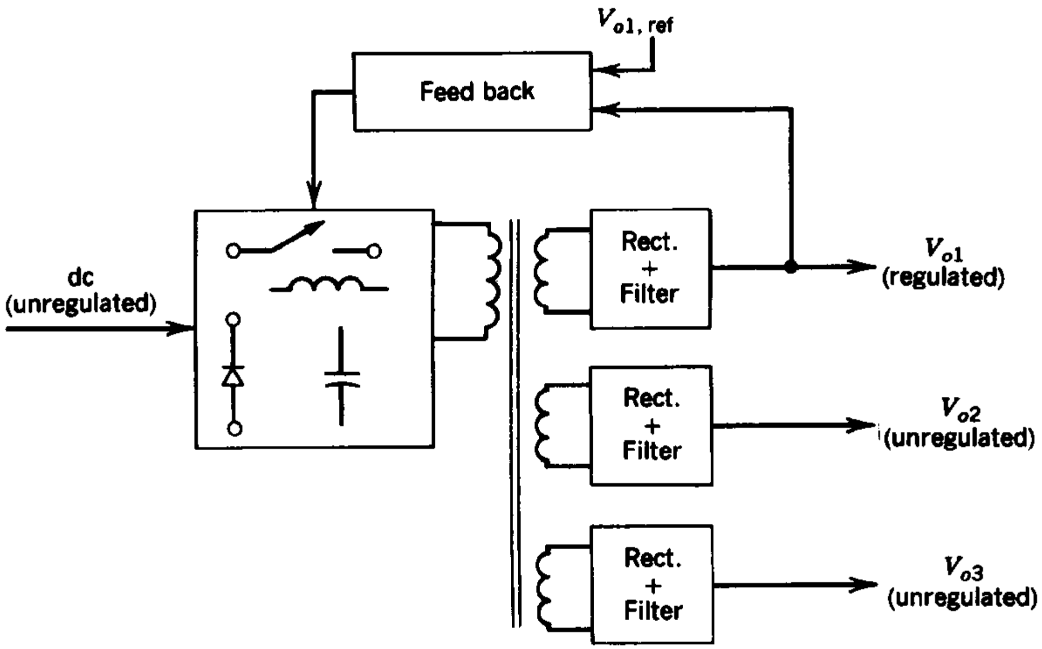
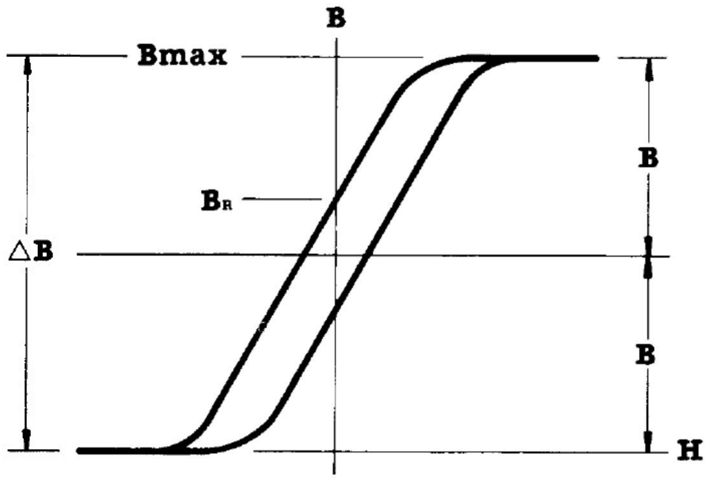


Figure 10-3 Multiple outputs.

Transformer core size

The power handling capacity of a transformer core can be determined by its WaAc product, where Wa is the available core window area, and Ac is the effective core cross-sectional area.



The WaAc/power-output relationship is obtained by starting with Faraday's Law:

$E = 4B\ Ac\ Nf \times 10^{-8}$ (square wave) (1)

$E = 4.44\ BAc\ Nf \times 10^{-8}$ (sine wave) (1a)

Where:

- E = applied voltage (rms)
- B = flux density in gauss
- Ac = core area in cm²
- N = number of turns
- f = frequency in Hz
- Aw = wire area in cm²
- Wa = window area in cm²:
  - core window for toroids
  - bobbin window for other cores
- C = current capacity in cm²/amp
- K = winding factor
- I = current (rms)
- P_i = input power
- P_o = output power
- e = transformer efficiency

Solving (1) for NAc

$$NAc = \frac{E \times 10^8}{4Bf} \qquad (2)$$

The winding factor

$$K = \frac{NAw}{Wa} \quad \text{thus } N = \frac{KWa}{Aw} \quad \text{and } NAc = \frac{KWaAc}{Aw} \qquad (3)$$

Combining (2) and (3) and solving for WaAc:

$$WaAc = \frac{E\ Aw \times 10^8}{4B\ fK}, \text{ where } WaAc = cm^4 \qquad (4)$$

In addition:

$C = Aw/I$ or $Aw = IC$ $e = P_o/P_i$ $P_i = EI$

Thus:

$E\ Aw = EIC = P_iC = P_oC/e$

Substituting for EAw in (4), we obtain:

$$\boxed{WaAc = \frac{P_oC \times 10^8}{4eB\ fK}} \qquad (5)$$

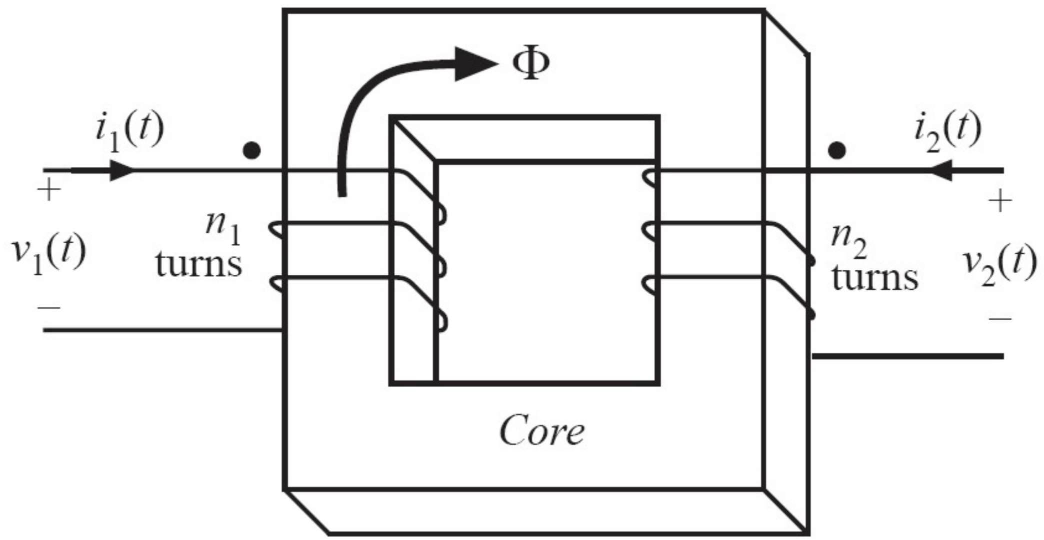
Transformer modeling

Two windings, no air gap:

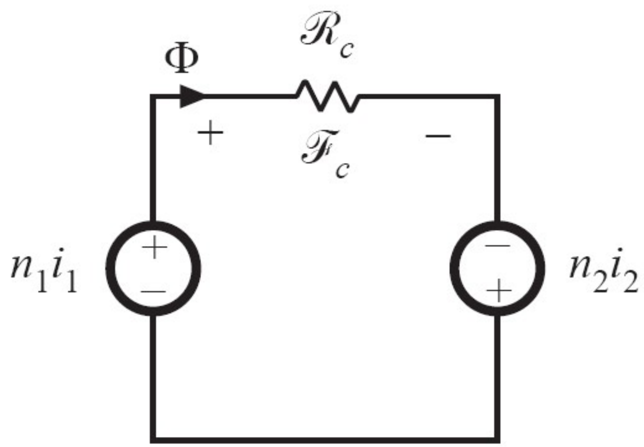
$$\mathcal{R}_c = \frac{\ell_m}{\mu A_c}$$

$$\mathcal{F}_c = n_1 i_1 + n_2 i_2$$

$$\Phi \mathcal{R}_c = n_1 i_1 + n_2 i_2$$



Magnetic circuit model:



Dot convention

The primary and secondary currents flowing into the winding terminals marked ● produce a mutually additive magnetic flux.

The ideal transformer

In the ideal transformer, the core reluctance $\mathscr{R}_c$ approaches zero.

MMF $\mathscr{F}_c = \Phi\, \mathscr{R}_c$ also approaches zero. We then obtain

$$0 = n_1 i_1 + n_2 i_2$$

Also, by Faraday's law,

$$v_1 = n_1 \frac{d\Phi}{dt}$$

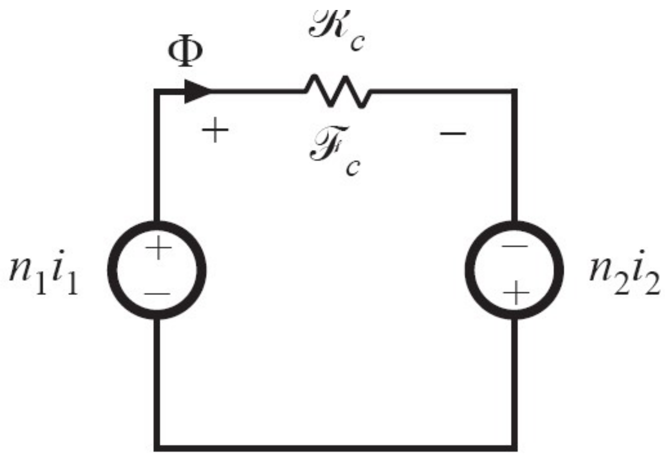
$$v_2 = n_2 \frac{d\Phi}{dt}$$

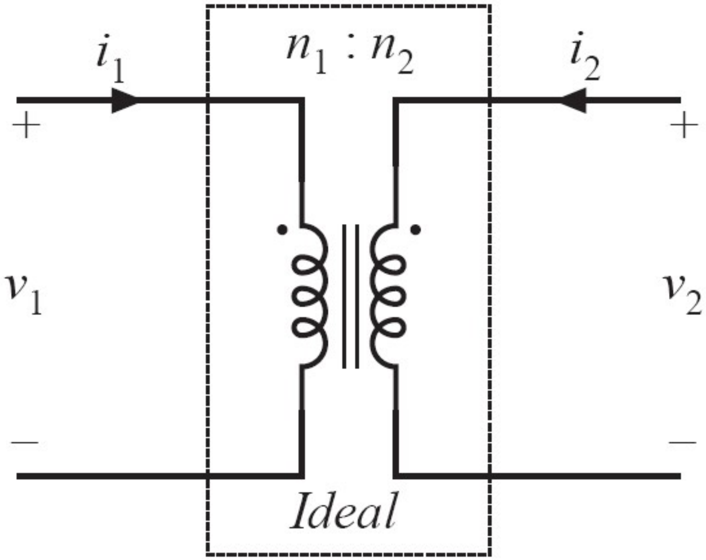
Eliminate Φ :

$$\frac{d\Phi}{dt} = \frac{v_1}{n_1} = \frac{v_2}{n_2}$$

Ideal transformer equations:

$$\frac{v_1}{n_1} = \frac{v_2}{n_2} \quad \text{and} \quad n_1 i_1 + n_2 i_2 = 0$$

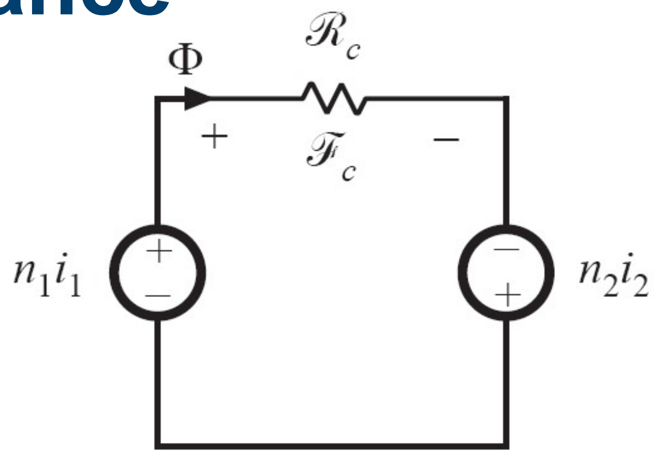




The magnetizing inductance

For nonzero core reluctance, we obtain

$$\Phi \mathcal{R}_c = n_1 i_1 + n_2 i_2 \quad \text{with} \quad v_1 = n_1 \frac{d\Phi}{dt}$$



Eliminate Φ:

$$v_1 = \frac{n_1^2}{\mathcal{R}_c} \frac{d}{dt}\left[i_1 + \frac{n_2}{n_1} i_2 \right]$$

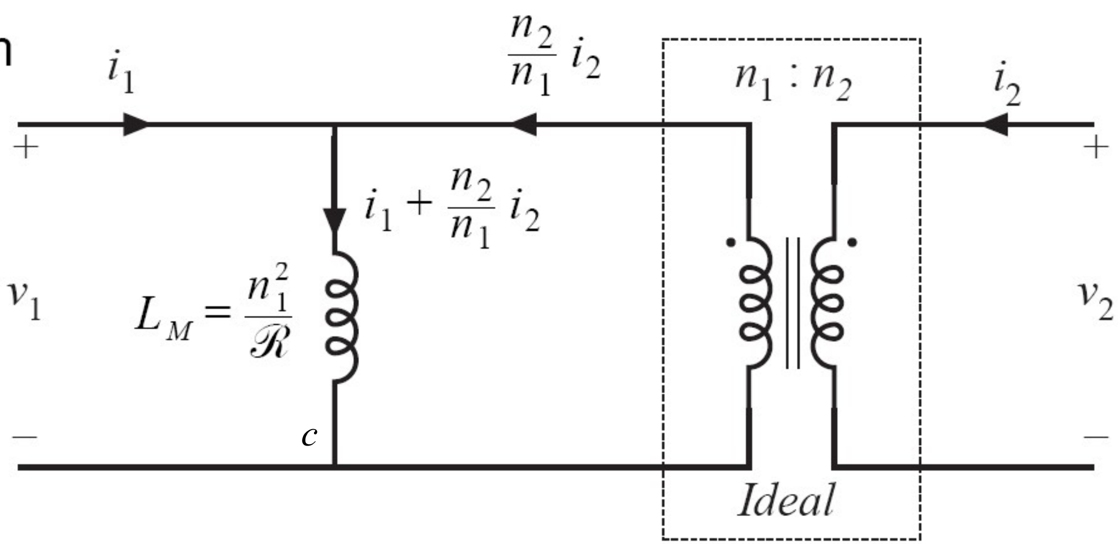
This equation is of the form

$$v_1 = L_M \frac{di_M}{dt}$$

with

$$L_M = \frac{n_1^2}{\mathcal{R}_c}$$

$$i_M = i_1 + \frac{n_2}{n_1} i_2$$



Magnetizing current $i_m(t)$: the current required to create $H_c(t)$ that couples primary and secondary windings through magnetic induction.

The magnetizing inductance (cont’d)

- Models magnetization of core material
- A real, physical inductor, that exhibits saturation and hysteresis
- If the secondary winding is disconnected:

  we are left with the primary winding on the core

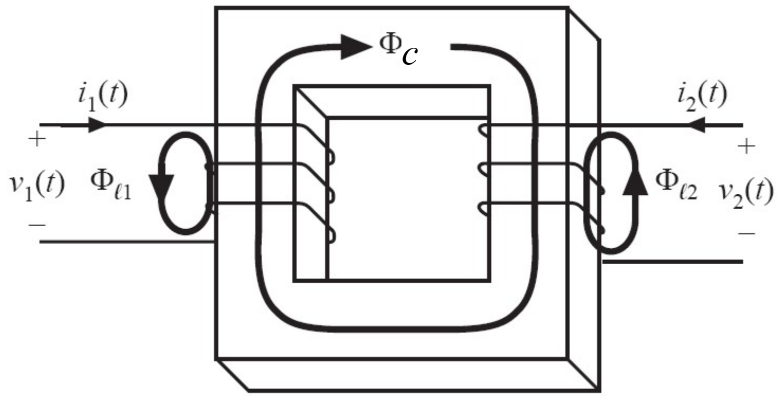
  primary winding then behaves as an inductor

  the resulting inductor is the magnetizing inductance, referred to the primary winding
- Magnetizing current causes the ratio of winding currents to differ from the turns ratio

Add leakage fluxes

$$\phi_c = \frac{N_1 i_1 + N_2 i_2}{\Re_c}; \quad \phi_{l1} = \frac{N_1 i_1}{\Re_{l1}}; \quad \phi_{l2} = \frac{N_2 i_2}{\Re_{l2}}$$

$$\lambda_1 = N_1\left(\phi_c + \phi_{l1}\right); \quad \lambda_2 = N_2\left(\phi_c + \phi_{l2}\right)$$



$$v_1 = \frac{d\lambda_1}{dt} = \boxed{L_{l1}\frac{di_1}{dt}} + L_m\left(\frac{di_1}{dt} + \frac{N_2}{N_1}\frac{di_2}{dt}\right) \quad (1)$$

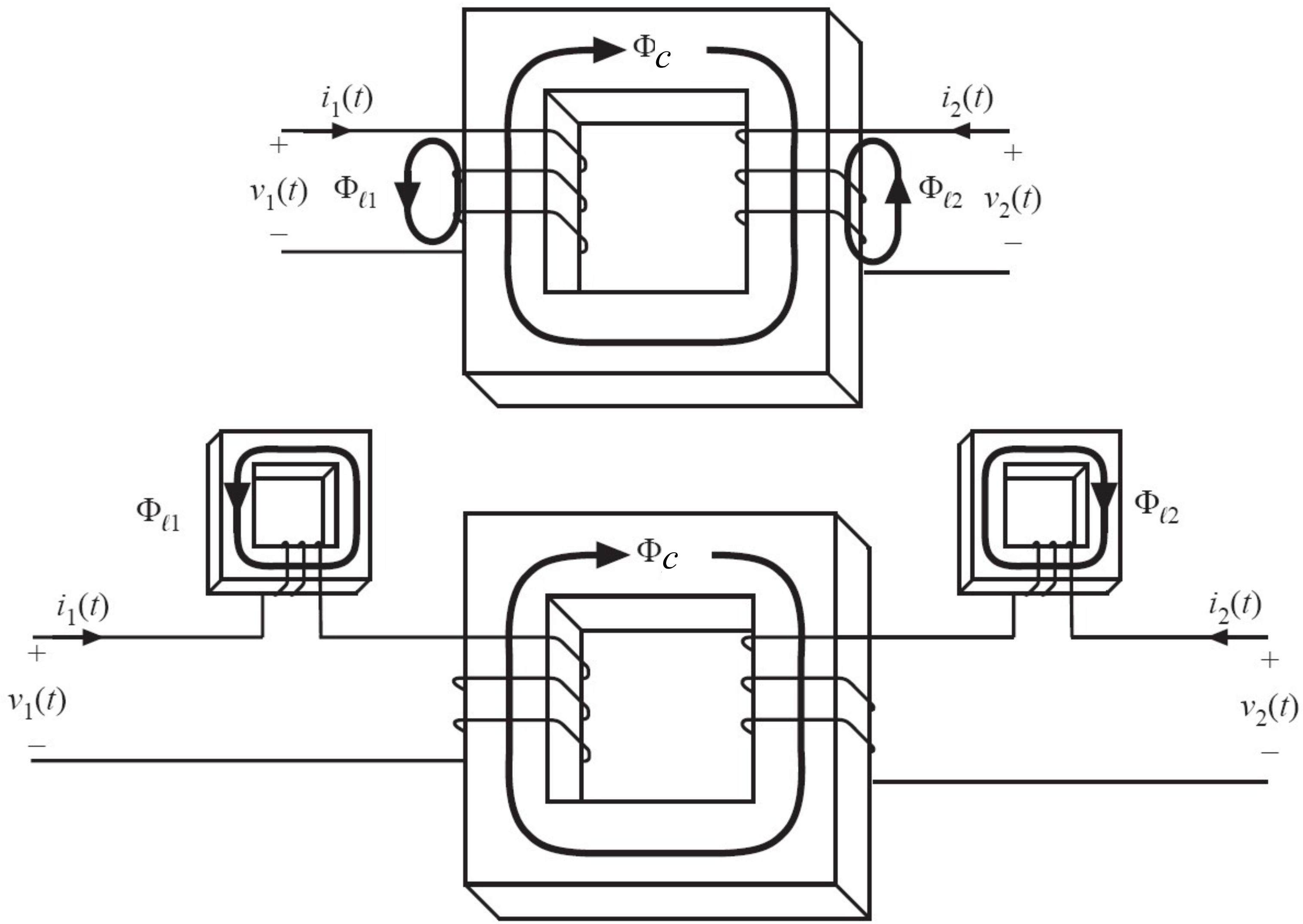
$$v_2 = \frac{d\lambda_2}{dt} = \boxed{L_{l2}\frac{di_2}{dt}} + \frac{N_2}{N_1}L_m\left(\frac{di_1}{dt} + \frac{N_2}{N_1}\frac{di_2}{dt}\right) \quad (2)$$

with

$$L_m = \frac{N_1^2}{\Re_c}; \quad L_{l1} = \frac{N_1^2}{\Re_{l1}}; \quad L_{l2} = \frac{N_2^2}{\Re_{l2}}$$

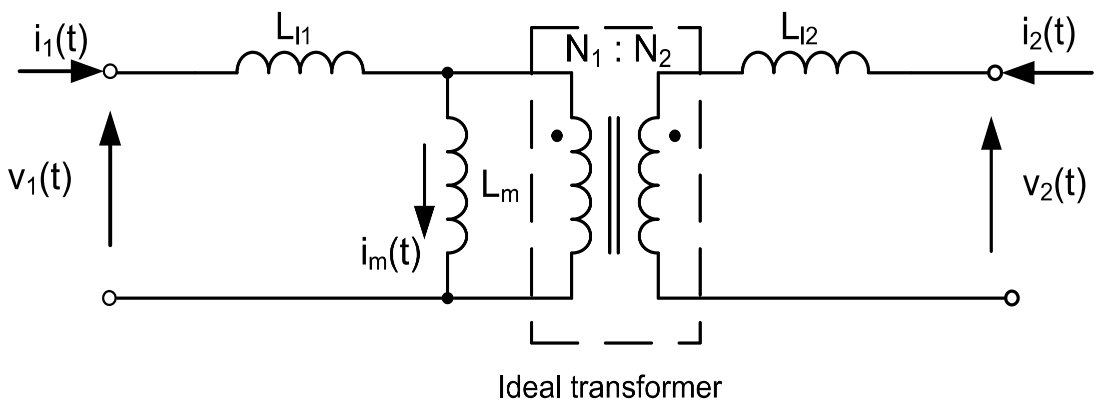
- Equations (1) and (2) provide a mathematical description of the input-output characteristics of a generic transformer

Leakage inductances



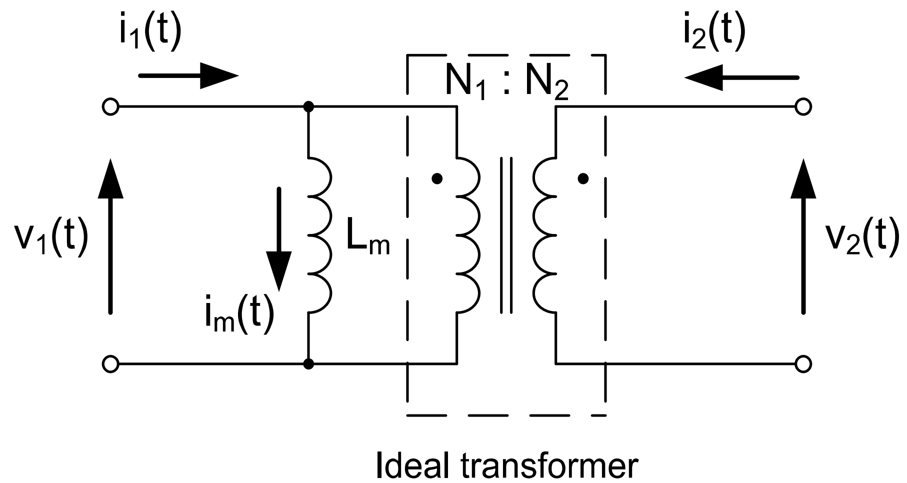
Transformer model (π - model)

An equivalent model that satisfies Eqs. (1) and (2) is illustrated below. This model is known as the π-model for a two-winding transformer.

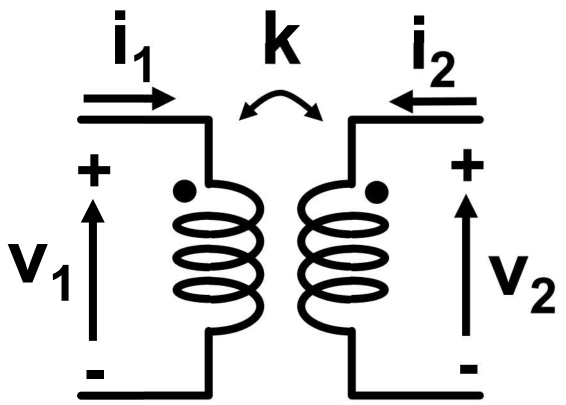


<u>Note</u>: core and copper losses are neglected

In the initial analysis we'll use this simplified model, assuming an almost ideal magnetic coupling ($k \approx 1$)



Transformer: coupled inductors model



$$\lambda_1 = L_1 i_1 + M i_2$$

$$\lambda_2 = M i_1 + L_2 i_2$$

$$v_1 = \frac{d\lambda_1}{dt} = L_1 \frac{di_1}{dt} + M \frac{di_2}{dt}$$

$$v_2 = \frac{d\lambda_2}{dt} = M \frac{di_1}{dt} + L_2 \frac{di_2}{dt}$$

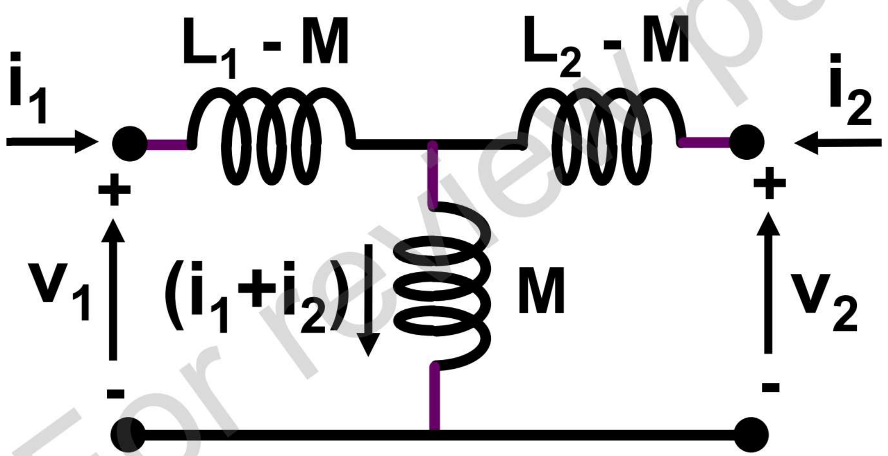
- Three parameters (L_1, L_2, M)
  - $L_1 = \frac{1}{\omega}\left(\frac{\hat{V}_1}{\hat{I}_1}\right)$
  - $L_2 = \frac{1}{\omega}\left(\frac{\hat{V}_2}{\hat{I}_2}\right)$
  - $M = \frac{1}{\omega}\left(\frac{\hat{V}_2}{\hat{I}_1}\right)$
- No direct influence of N_1, N_2 evident

Note: coupling factor $k = \frac{M}{\sqrt{L_1 L_2}}$

Transformer equivalent circuit

$$v_1 = L_1 \frac{di_1}{dt} + M \frac{di_2}{dt} = (L_1 - M)\frac{di_1}{dt} + M\frac{d}{dt}(i_1 + i_2)$$

$$v_2 = M \frac{di_1}{dt} + L_2 \frac{di_2}{dt} = M\frac{d}{dt}(i_1 + i_2) + (L_2 - M)\frac{di_2}{dt}$$

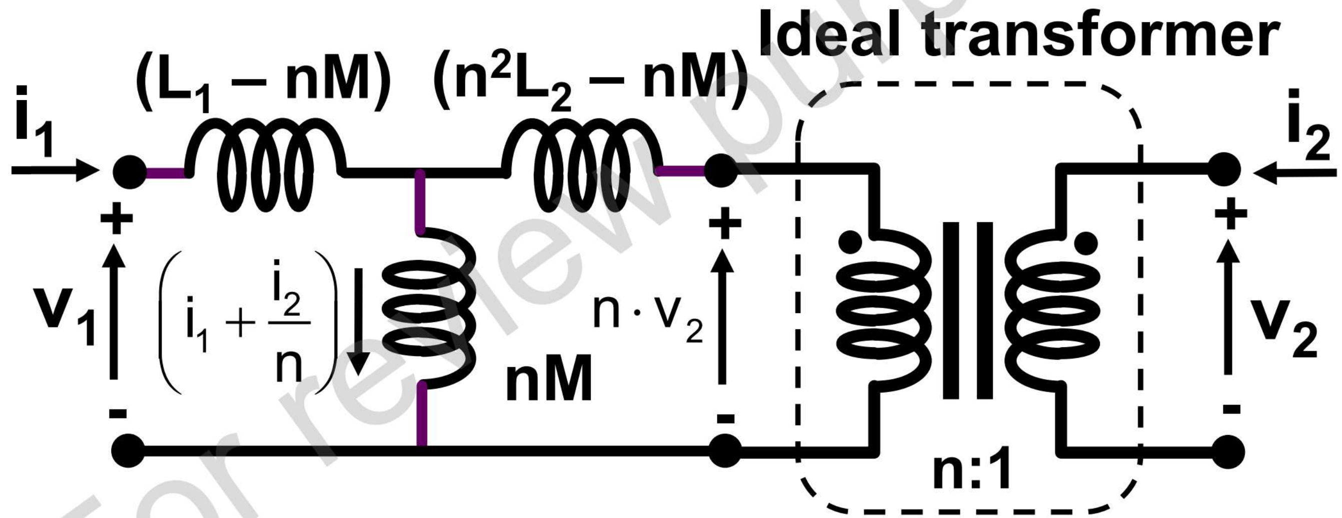


T - model

Introduction of a general transformation ratio n

$$v_1 = L_1\frac{di_1}{dt} + M\frac{di_2}{dt} = (L_1 - nM)\frac{di_1}{dt} + nM\frac{d}{dt}\left(i_1 + \frac{i_2}{n}\right)$$

$$v_2 = M\frac{di_1}{dt} + L_2\frac{di_2}{dt} = \frac{1}{n}\left(nM\frac{d}{dt}\left(i_1 + \frac{i_2}{n}\right) + (n^2L_2 - nM)\frac{d}{dt}\left(\frac{i_2}{n}\right)\right)$$

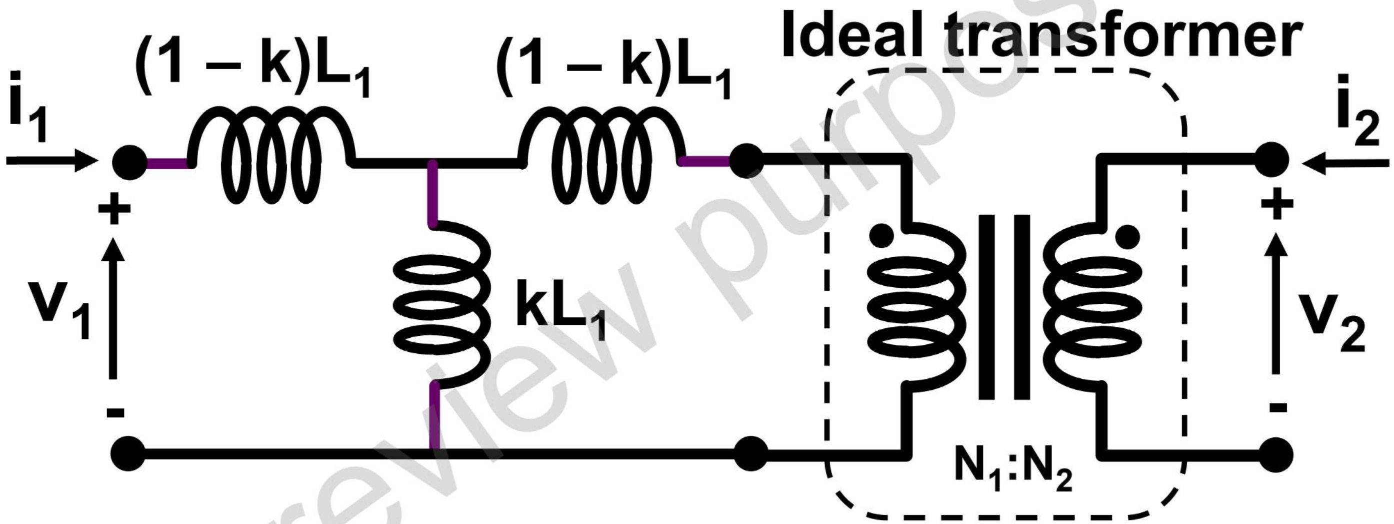


Note: the parameter n can be arbitrarily chosen, generating an infinite array of circuit models with the same terminal characteristics!

Choice #1

$$n = \sqrt{\frac{L_1}{L_2}} \approx \frac{N_1}{N_2} \quad \text{if } k \text{ is close to } 1$$

With this choice, n is almost equal to the physical turn ratio



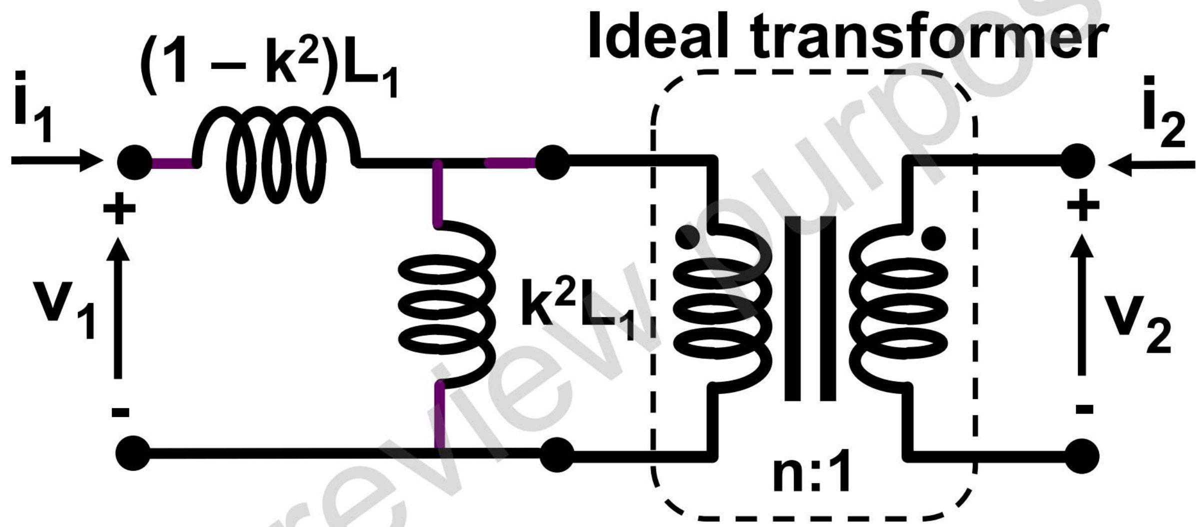
- Similar to the π - model (L_{l2} transferred to the primary side) with L_m=kL_1, L_{l1}=$(1-k)L_1$, L_{l2}=$(1-k)L_2$

- Note: $v_2/v_1 \approx k\, N_2/N_1$ with secondary-side terminals open

Choice #2

$$n = k \cdot \sqrt{\frac{L_1}{L_2}}$$

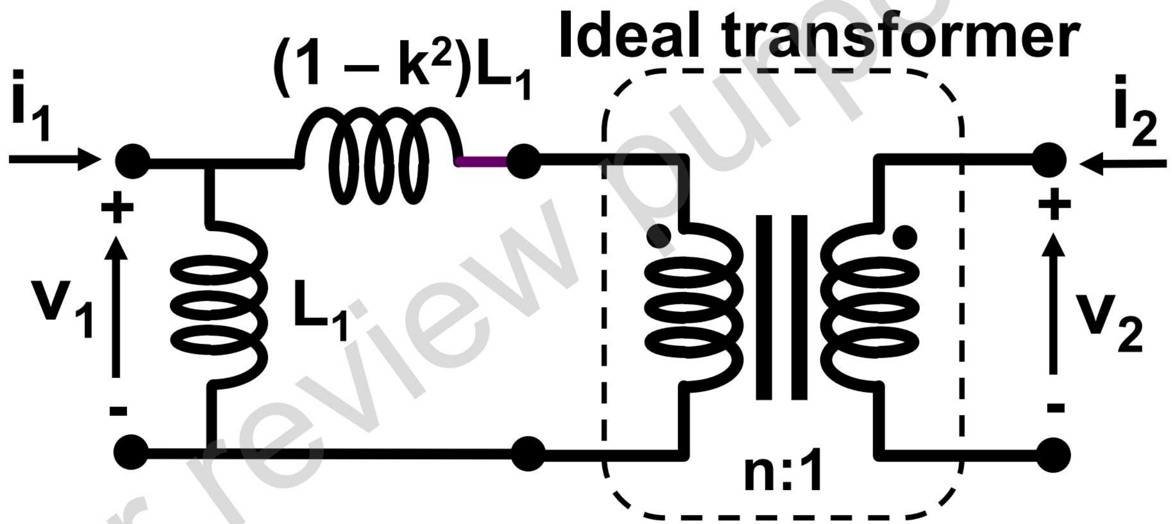
→ This choice nulls the secondary leakage inductance



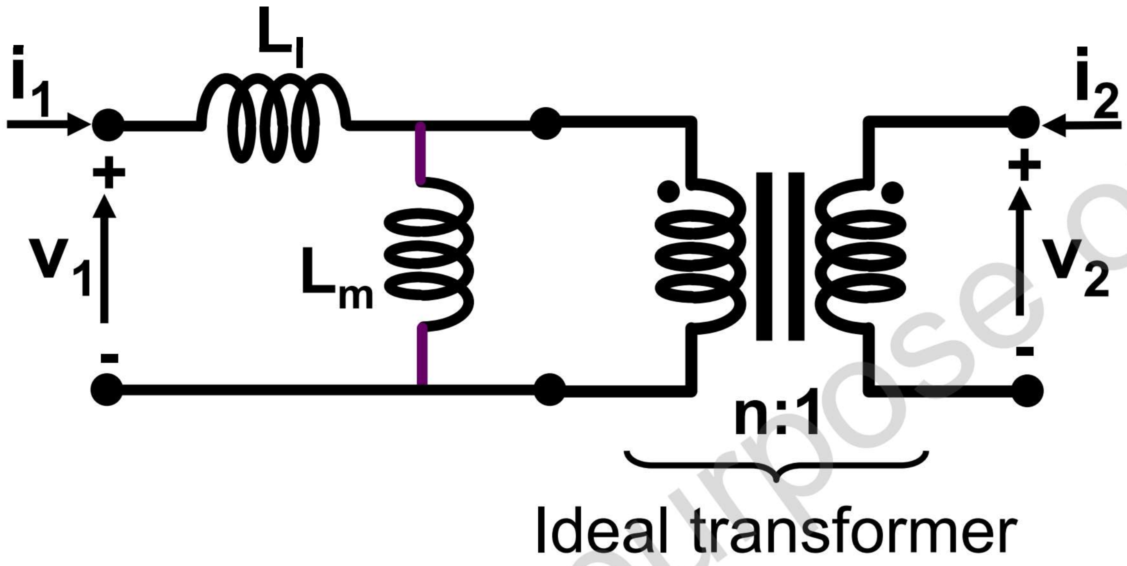
Choice #3

$$n = \frac{1}{k} \cdot \sqrt{\frac{L_1}{L_2}}$$

→ This choice nulls the primary leakage inductance

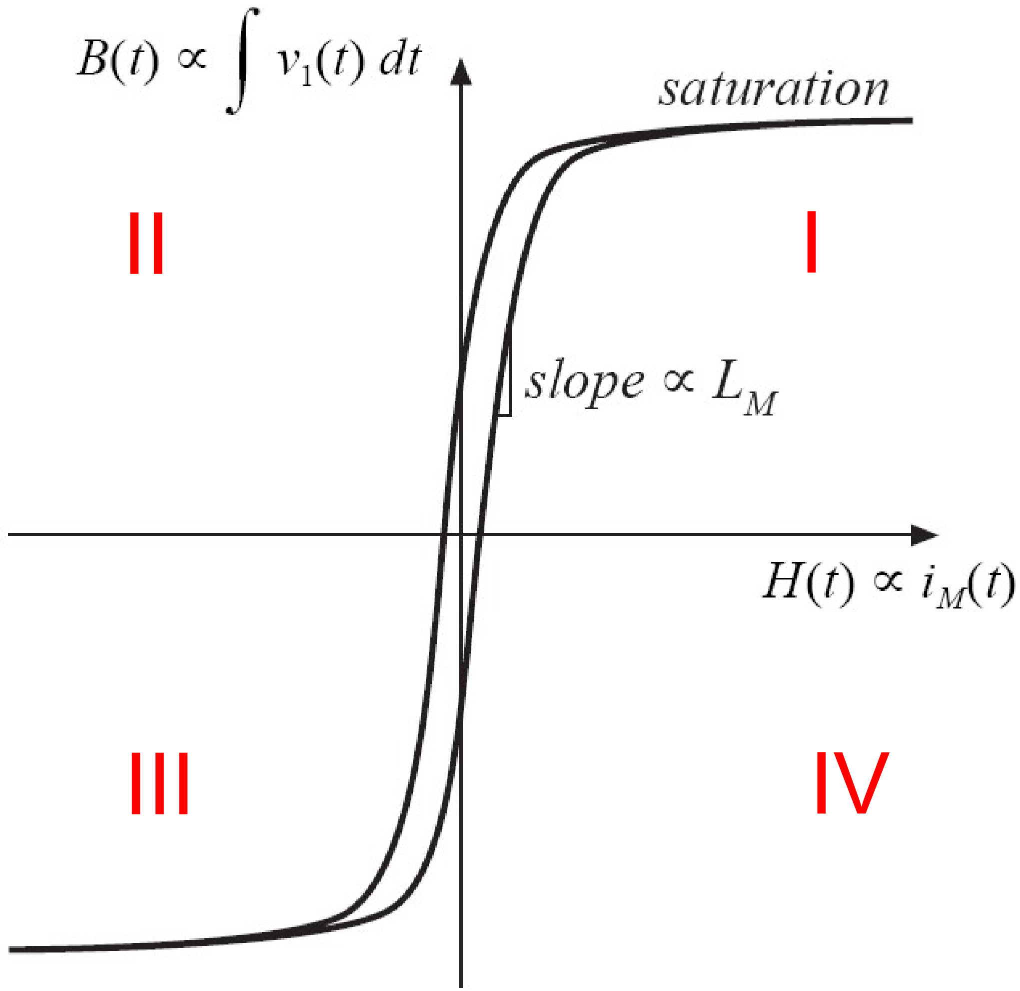


Commonly used model (k close to 1)



- L_l is obtained by measuring the primary-side inductance with the secondary-side terminals shorted
- L_m is obtained by measuring the primary-side inductance with the secondary-side terminals open.
  - ➢ $L_m = L_{measured} - L_l$
- n is achieved by measuring the primary side voltage (open terminals) upon the application of a voltage to the secondary side

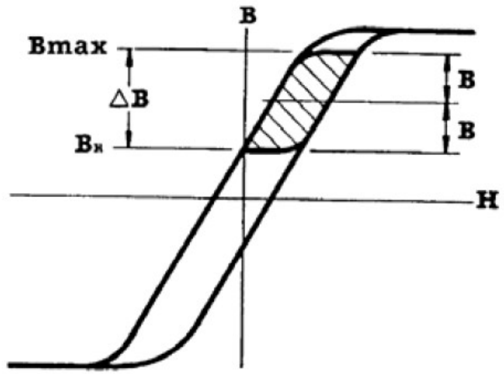
Transformer core B-H characteristics



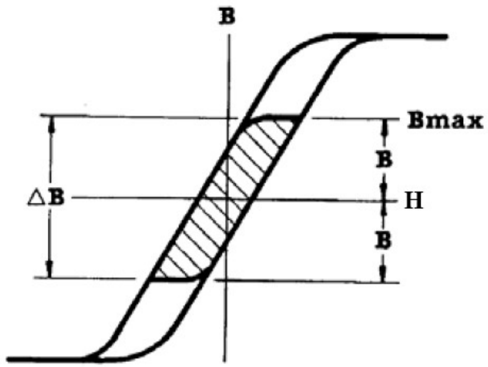
Isolated converter topologies /1

Two kinds of core excitation for the above converters

- Unidirectional core excitation, where only the positive part (quadrant I) of the B-H loop is used:
  - flyback coverter (derived from buck-boost converter)
  - forward converter (derived from buck converter).
- Bidirectional core excitation, where the positive (quadrant I) and the negative (quadrant III) parts of the B-H loop are utilized alternatively
  - push-pull converter
  - half-bridge converter
  - full-bridge converters

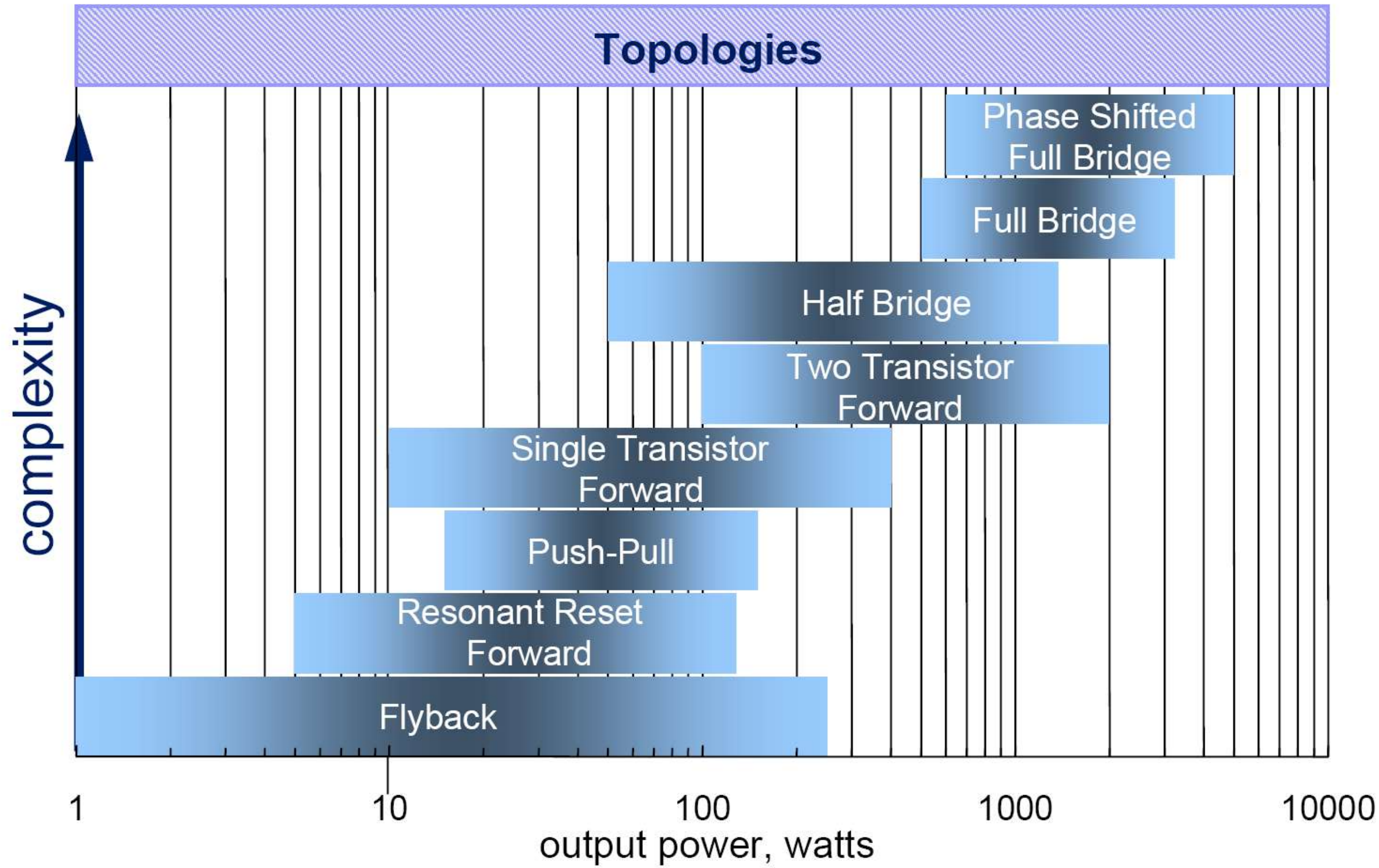


"Single-Ended"
AC operation with DC offset



"Double-ended"
AC-only Operation

Isolated converter topologies /2



Flyback converter

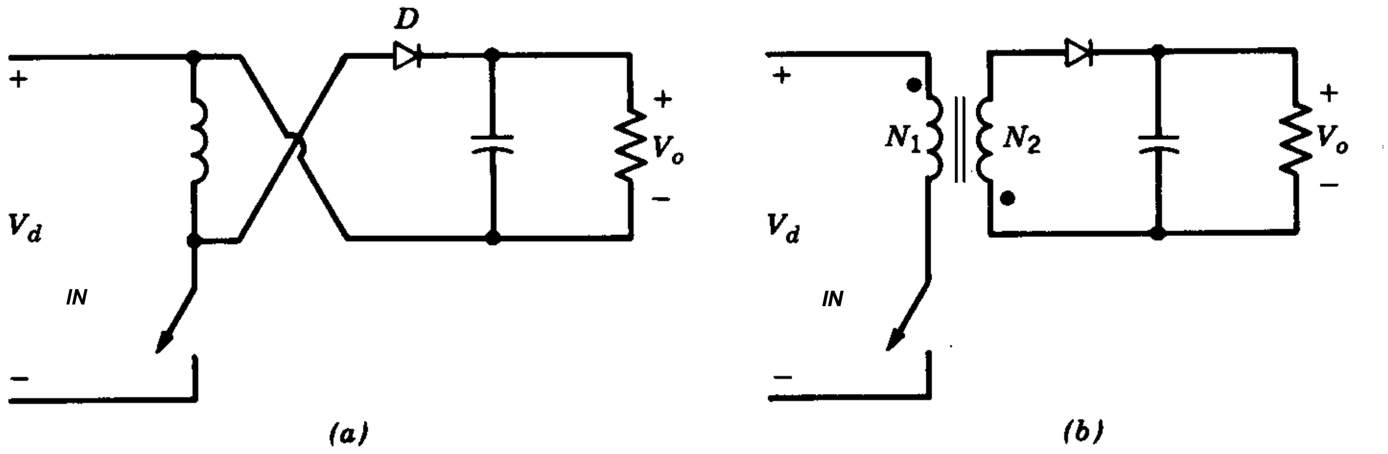


Figure 10-6 Flyback converter.

- Flyback is the most popular topology for low-power (<100 W) AC/DC conversion applications:
- a flyback transformer combines the actions of an isolating transformer and an output inductor into a single element hence, no separate inductor is needed;
- a flyback transformer provides galvanic (electrical) isolation up to ~5 kV DC, and it can provide multiple and/or negative outputs;
- simple design requires just one semiconductor switch (MOSFET) and one freewheeling diode.

Flyback converter: CCM

- During t_{on}, V_{in} is applied to primary and current ramps up in L_m.

- As switch opens, voltage "flies back" until caught by secondary diode / cap at V_{out} on secondary side, $= (-N_1/N_2\ V_{out})$ on primary winding (L_m).

- During t_{off}, current ramps down in L_m.

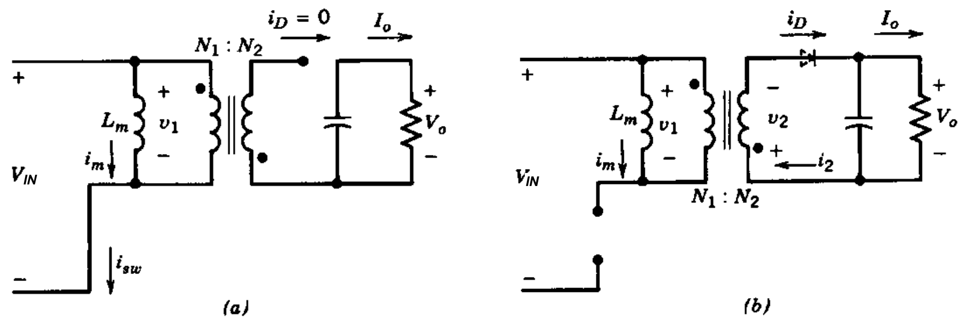
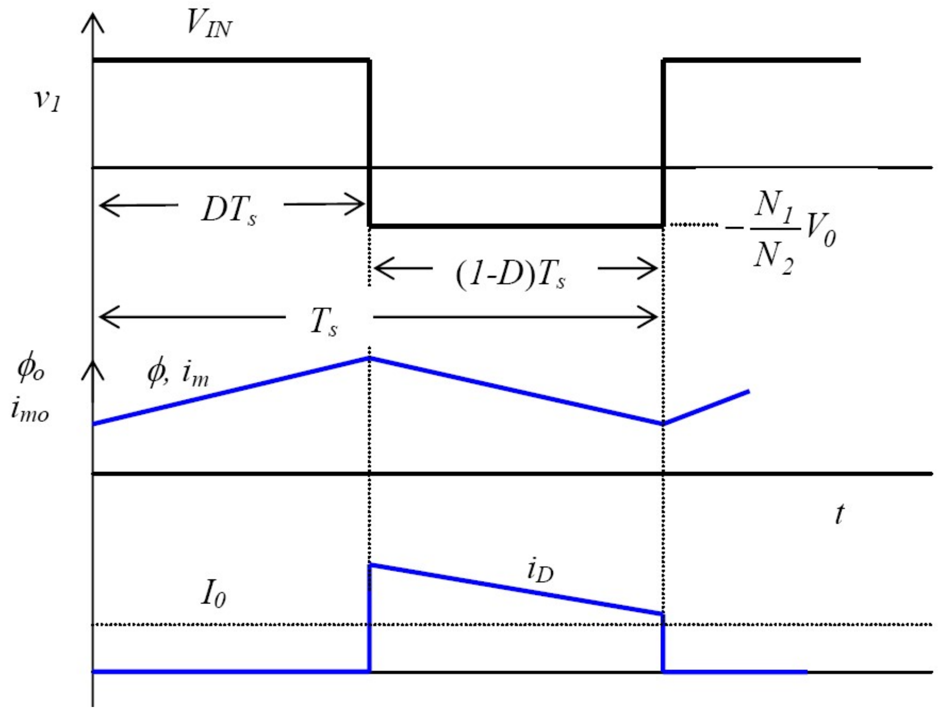
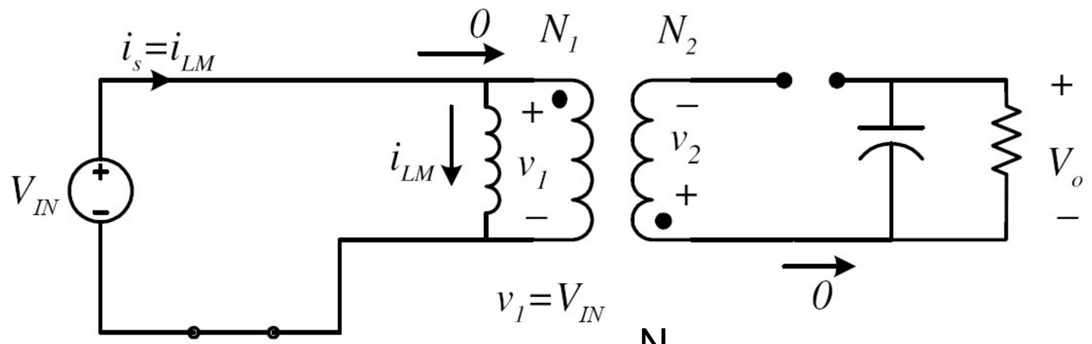


Figure 10-7 Flyback converter circuit states: (*a*) switch on; (*b*) switch off.



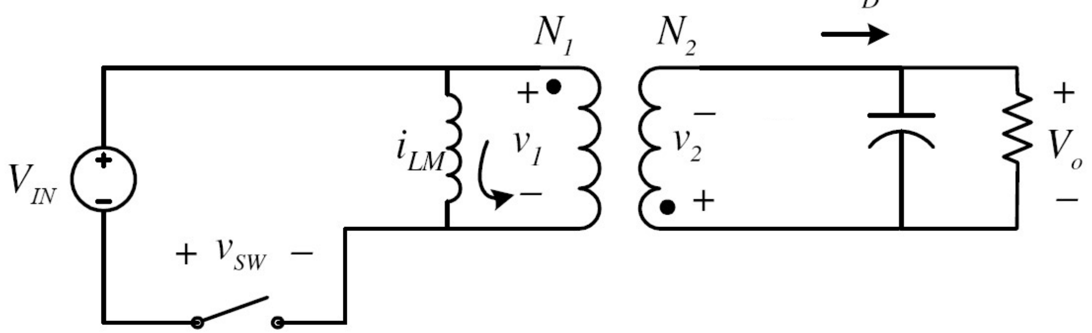
Flyback converter: CCM



$v_1 = V_{IN}$

$v_2 = V_{IN}\dfrac{N_2}{N_1}$

$$\Delta i^{+}_{L_M} = \frac{V_{IN} D T_S}{L_M}$$



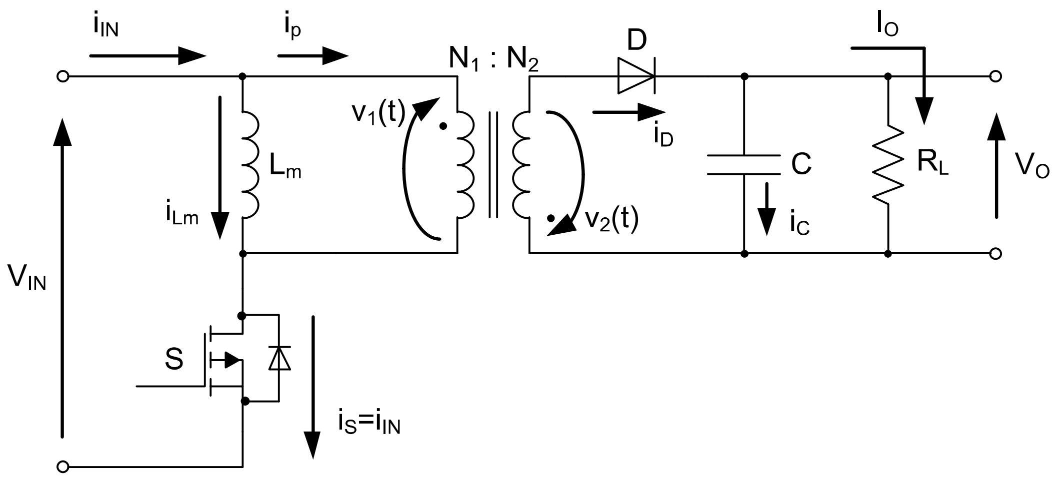
$$\Delta i^{-}_{L_M} = \frac{N_1}{N_2}\frac{V_O(1-D)T_S}{L_M}$$

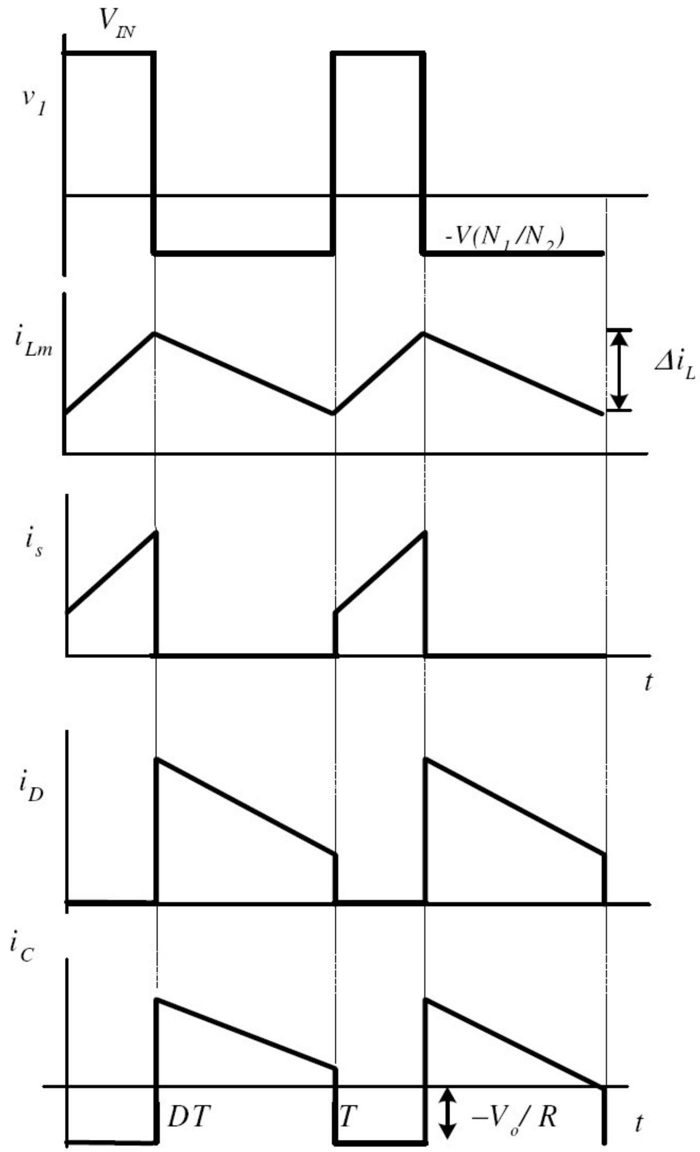
OFF

$$\Delta i^{+}_{L_M} = \Delta i^{-}_{L_M} \Rightarrow \boxed{\frac{V_O}{V_{IN}} = \frac{N_2}{N_1}\frac{D}{(1-D)}}$$

Input output relationship is similar to buck-boost converter.

Flyback converter: CCM waveforms





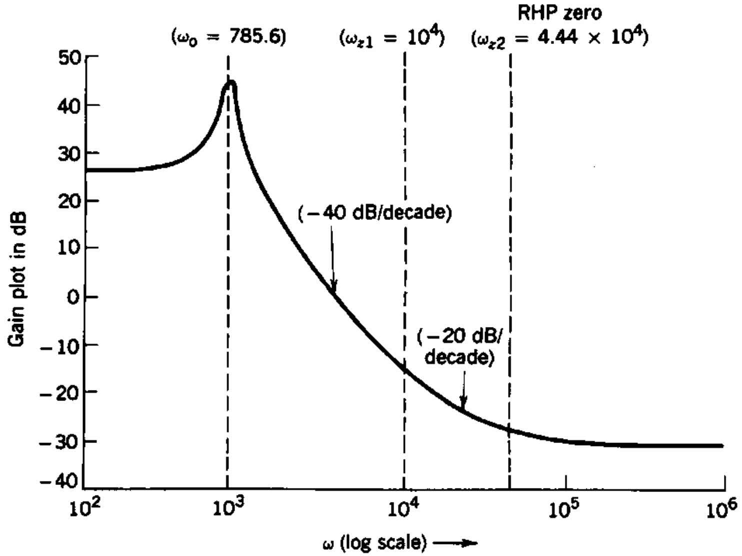
Flyback transformer

- **Conventional transformers store minimal energy**
  - E.g. Forward converter, Push-Pull converter etc.
  - Primary and secondary currents flow simultaneously.
  - Energy is transferred by the simultaneous flow of current.
- **Flyback "transformers" are really coupled-inductors**
  - Primary and secondary windings do not conduct simultaneously → they can be seen as two magnetically coupled inductors → often called inductor-transformer.
  - Current flows in primary while secondary diode is reverse-biased.
  - Primary current stores energy in air-gap.
  - When primary current stops, secondary winding reverses polarity □ forward-biases output diode → secondary current flows.
  - Air-gap energy is transferred to the secondary load:
    - **CCM** => only some of the stored energy is delivered to the load, primary current starts before secondary current has decayed to zero;
    - **DCM** => all stored energy is delivered to the load, followed by an interval of zero current flow in both windings.

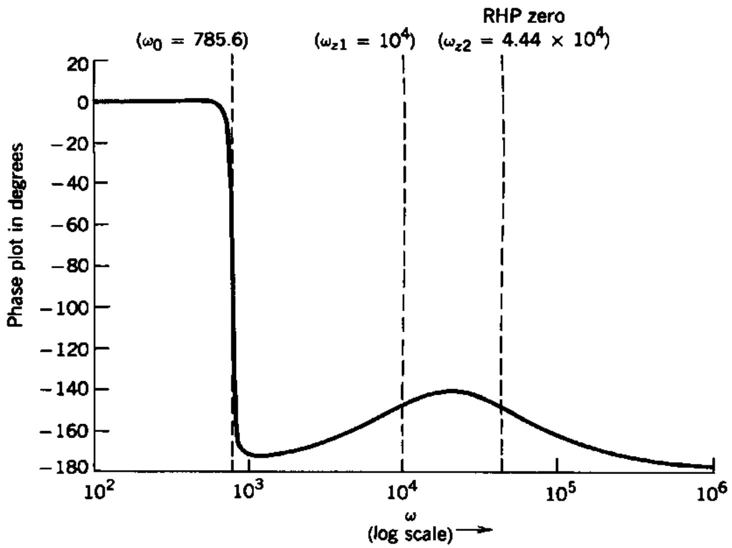
CCM flyback converter: RHP zero

$$T_{od}(s) = \frac{\tilde{v}_O(s)}{\tilde{d}(s)} = T_{od}(0)\frac{\left(1+\dfrac{s}{2\pi f_{z1}}\right)\left(1-\dfrac{s}{2\pi f_{z2}}\right)}{1+\dfrac{s}{2\pi f_O Q}+\left(\dfrac{s}{2\pi f_O}\right)^2}$$

A right half plane (RHP) zero is produced in the duty cycle-to-output transfer function when the flyback is operated in CCM.



gain

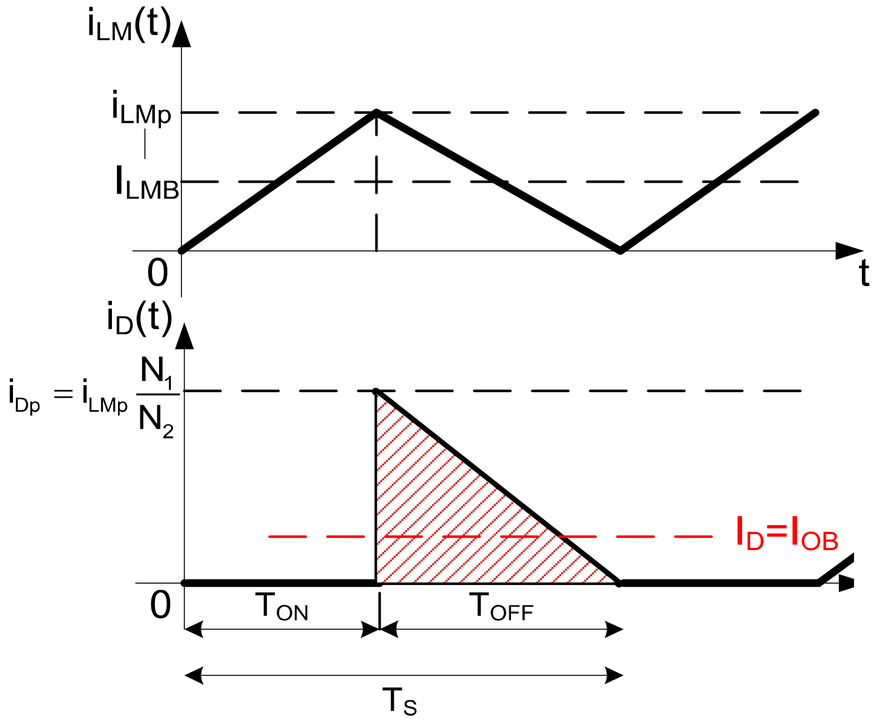


phase

Flyback converter: CCM-DCM boundary

$$I_{LMB} = \frac{i_{LMp}}{2} = \frac{V_{IN} D T_S}{L_m}$$

$$I_{OB} = I_D = i_{LMp} \cdot \frac{N_1}{N_2} \frac{(1-D)T_S}{2T_S} = \frac{V_{IN} D(1-D)}{2 L_m f_S} \cdot \frac{N_1}{N_2}$$



Using:

$$\frac{V_O}{V_{IN}} = \frac{N_2}{N_1} \frac{D}{(1-D)}$$

we get:

$$\boxed{I_{OB} = \frac{V_O \cdot (1-D)^2}{2 L_m f_S} \left(\frac{N_1}{N_2}\right)^2}$$

and:

$$\boxed{P_{OB} = \frac{V_O^2 \cdot (1-D)^2}{2 L_m f_S} \left(\frac{N_1}{N_2}\right)^2 = \frac{V_{IN}^2 D^2}{2 L_m f_S}}$$

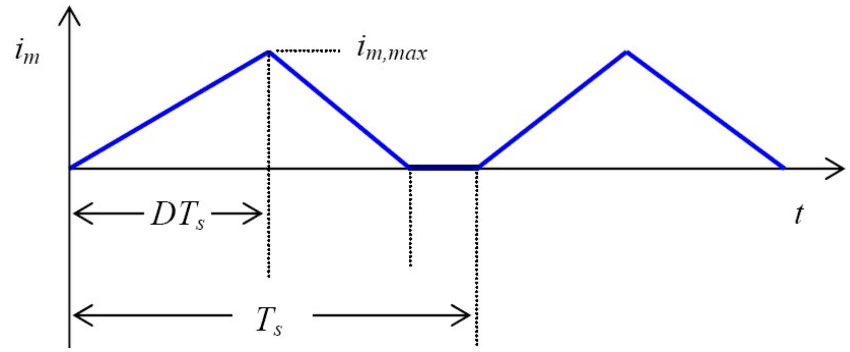
$\xrightarrow{\text{CCM}}$ $P_O > P_{OB}$

$\xrightarrow{\text{DCM}}$ $P_O < P_{OB}$

Flyback converter: DCM

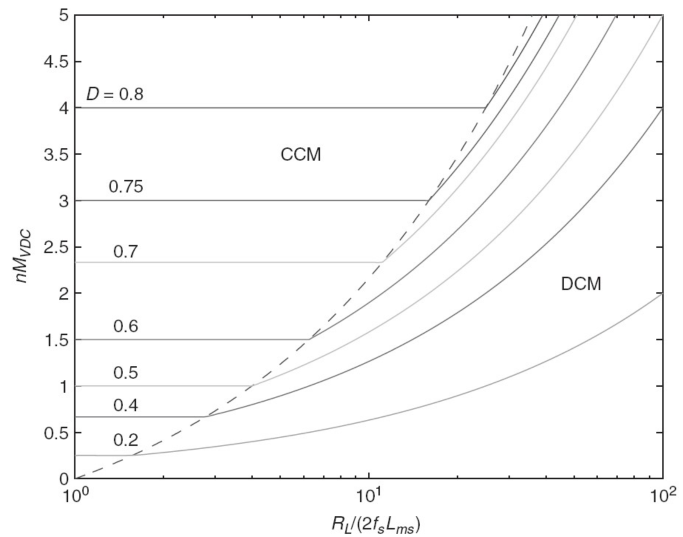
The energy transferred from the input dc voltage source V_{IN} to the magnetizing inductance during one cycle for the DCM case is:

$$W = \frac{1}{2} \cdot L_m \cdot i^2_{m,max}$$



Assuming 100% efficiency, this results in dc output power given by:

$$P_O = \frac{1}{2} \cdot L_m \cdot i^2_{m,max} \cdot f_S = \frac{V_O^2}{R}$$



since

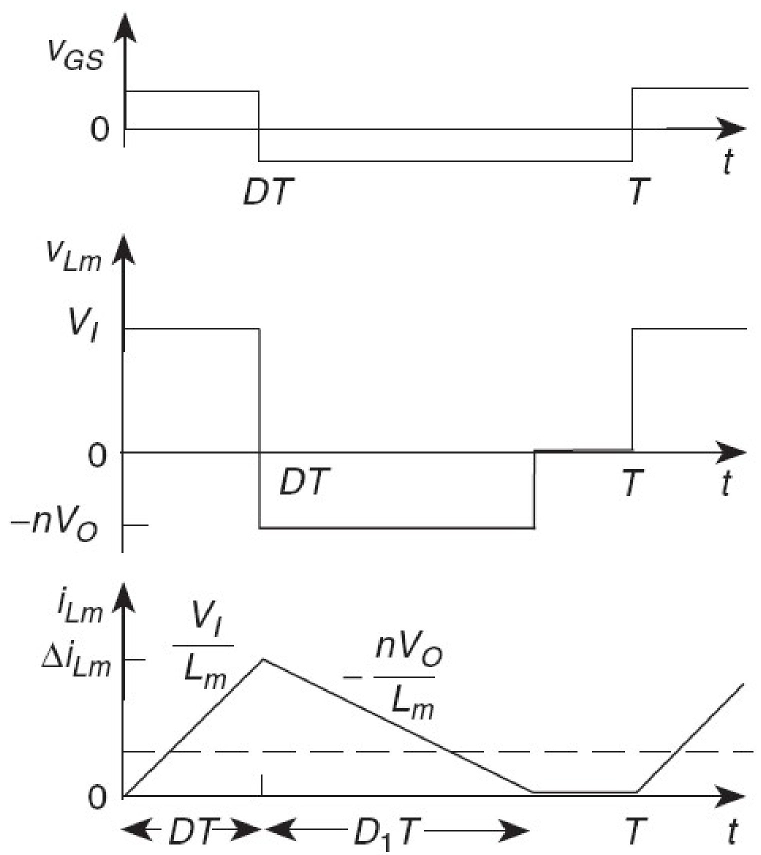
$$i_{m,max} = \left|\Delta i_m^+\right| = \frac{V_{IN} D}{L_m f_S}$$

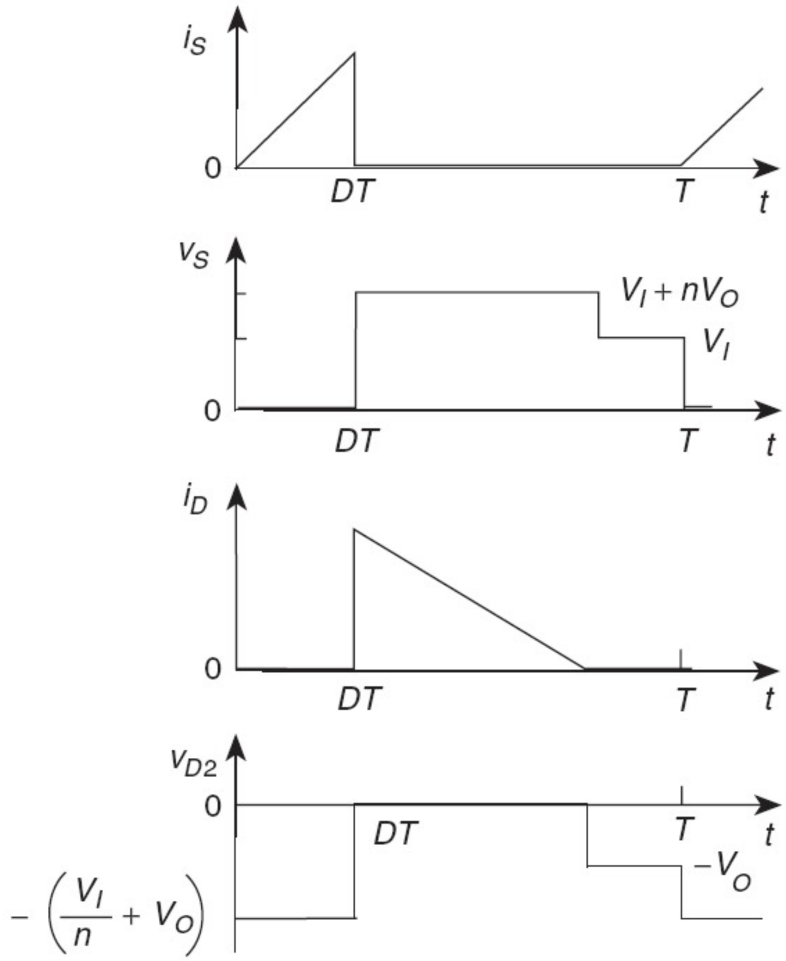
it follows that

$$\boxed{\frac{V_O}{V_{IN}} = D\sqrt{\frac{R}{2L_m f_S}}}$$

Thus the feedback loop will regulate the output by decreasing D as V_{IN} or R goes up, increasing D as V_{IN} or R goes down.

Flyback converter: DCM waveforms





$$\boxed{n = \frac{N_1}{N_2}}$$

Flyback converter: DCM vs CCM waveforms

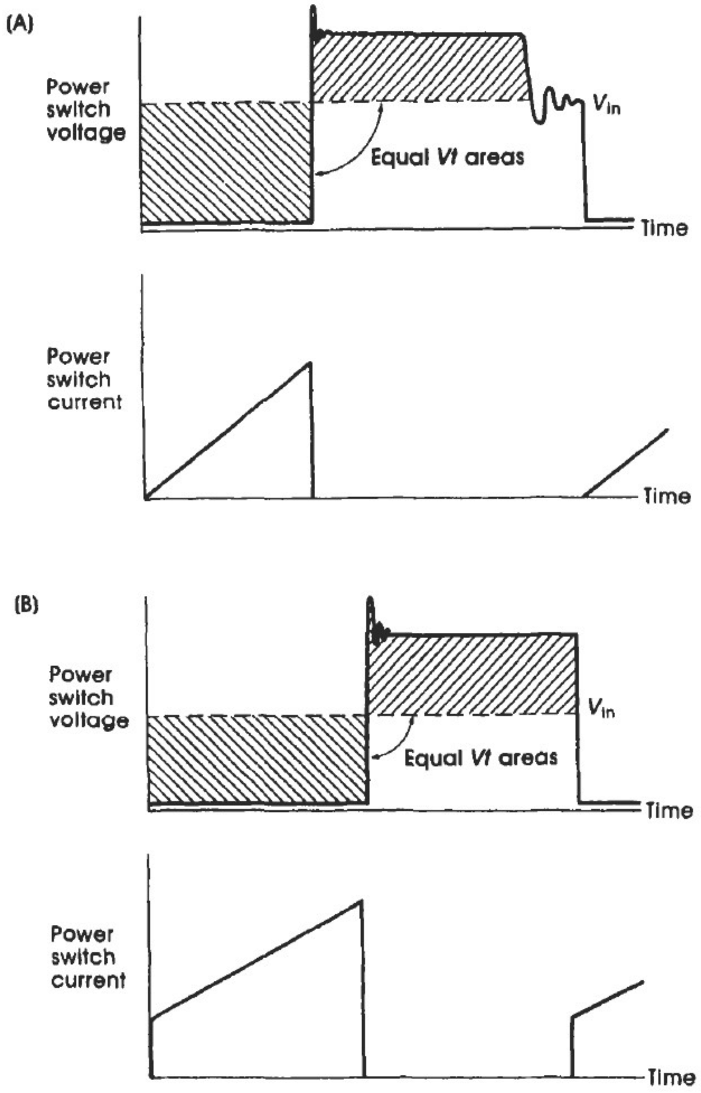


Figure 4.6
(A) Flyback operating in the discontinuous mode; (B) flyback operating in the continuous mode.

Challenges in flyback design

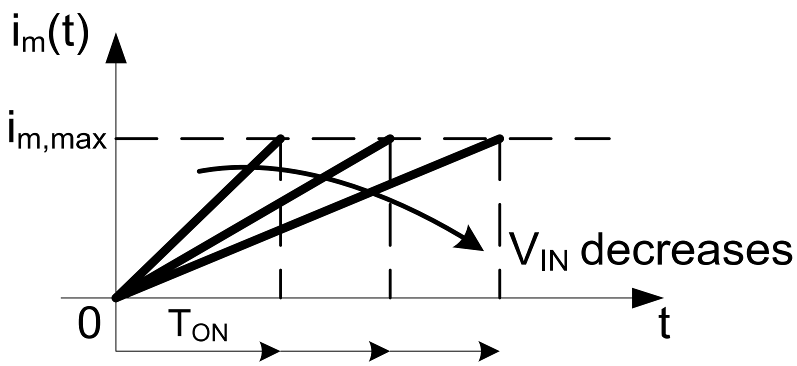
Overall Design	Transformer Design
• Balance cost, footprint and performance • Design decisions on modes of operation (CCM, DCM, CRM/TM) • Effective component selection – MOSFET, diodes should withstand the stress • Design RC or RCD snubber to suppress ringing • More complex design process compared to non-isolated topologies	• Customize for the specific power design specifications • Optimize selection of transformer based on efficiency, cost, footprint • Select appropriate core, considering loss and magnetic saturation • Select bobbin based on physical fit • Select wires gauge based on skin effect & current carrying capability • Design the winding strategy

DCM flyback converter: L_m design

- The magnetizing inductance L_m needs to be limited to a maximum value.
- DCM ensured if:

$$P_O < P_{OB} = \frac{V_{IN}^2 D^2}{2 L_m f_S} \quad \text{i.e.} \quad \boxed{L_m < \frac{V_{IN}^2 D^2}{2 P_O f_S}}$$

Worst conditions? V_{IN}? D? P_O?



- At a given P_O,

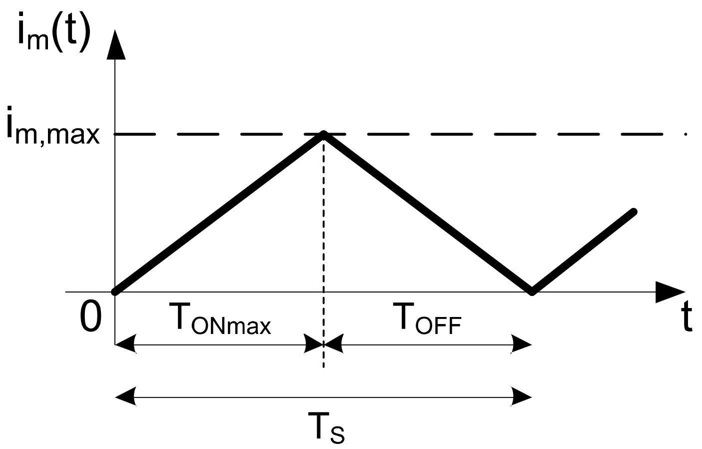
$$i_{m,max} = \text{const} = \sqrt{\frac{2 P_O}{L_m f_S}}$$

- The lower $V_{IN,\,MIN}$, the longer the T_{ON} needed to reach $i_{m,max}$

DCM flyback converter: L_m design

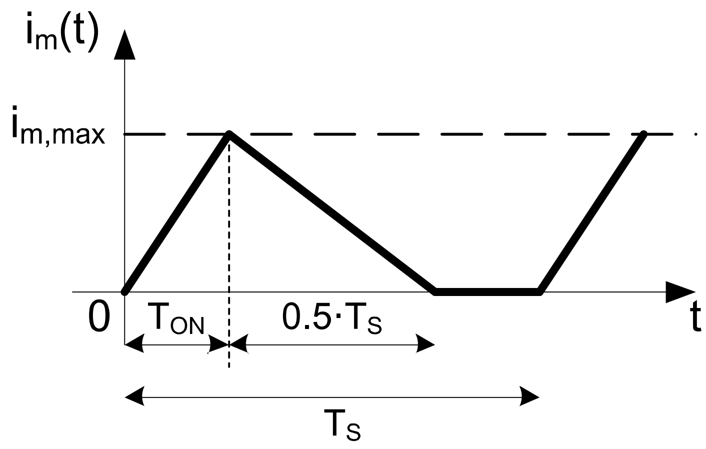
Guidelines

- The maximum T_{ON}, T_{ONmax} is usually set at $0.5T_S$ (i.e. D_{max}=50%)
- L_m is designed such that i_m is able to reach $i_{m,max}$ within T_{ONmax} in the worst conditions (P_{Omax}, V_{INmin})
- N_1/N_2 is designed such that the L_M is discharged within (T_S-$T_{ON,max}$)



$V_{IN} = V_{INmin}$

$T_{ON} = T_{ONmax} = 0.5\ T_S$



$V_{IN} > V_{INmin}$

$T_{ON} < T_{ONmax}$

DCM flyback converter: L_m design

$$\boxed{L_m \le \frac{V_{INmin}^2 \cdot D_{max}^2}{2\,P_{Omax} \cdot f_S}} \quad \xrightarrow{\text{Assuming } D_{max}=50\%} \quad \boxed{L_m \le \frac{V_{INmin}^2}{8\,P_{Omax} \cdot f_S}}$$

If converter efficiency, η, is taken into account, we easily get:

$$\boxed{L_m \le \frac{\eta \cdot V_{INmin}^2}{8\,P_{Omax} \cdot f_S}}$$

- Assume η ~ 0.75 - 0.85 as a first guess

DCM flyback converter: turn ratio design

$$\Delta i^{+}_{L_M} = \Delta i_{L_M} \Longrightarrow \frac{V_O}{V_{IN}} = \frac{N_2}{N_1}\frac{D}{(1-D)}$$

For $D=D_{max}=0.5$, $V_{IN}=V_{INmin}$ (worst conditions) we get:

$$\boxed{\frac{N_1}{N_2} = \frac{V_{INmin}}{V_O}}$$

More accurately, taking into account non-idealities:

$$\boxed{\frac{N_1}{N_2} = \left(\frac{V_{INmin} - V_{DSon}}{V_O + V_D}\right)}$$

Flyback regulator: design example

- Input voltage: $V_{IN} = 36 \div 72$ V
- Output voltage: $V_O = 5$ V
- DCM operation
- Output voltage ripple: $\Delta V_O = 100$ mV
- Output current: $I_O = 8$ A $\rightarrow P_o = 40$ W
- Switching frequency: $f_S = 100$ KHz
- Closed loop bandwidth: $f_C = 20$ kHz

Flyback regulator: design example

Step 1: determine the turn ratio

$$\frac{N_1}{N_2} = \left(\frac{V_{INmin} - V_{DSon}}{V_O + V_D} \right) \approx 6$$

Assume V_d and V_{DSon} ~ 1V

Step 2: determine switch voltage rating

$$BV_{DSmin} > V_{INmax} + \frac{N_1}{N_2} \cdot V_O = 102 \text{ V}$$

Power MOS

$$BV_D > \frac{N_2}{N_1} \cdot V_{INmax} + V_O = 17 \text{ V}$$

diode

Flyback regulator: design example

Step 3: determine the magnetizing inductance

$$L_m \le \frac{\eta \cdot \left(V_{INmin} - 1\right)^2}{8\, P_O \cdot f_S} \approx 30\,\mu H \qquad \text{(assume } \eta = 0.8\text{)}$$

Step 4: determine the peak primary current in order to fully specify the power switch

$$i_{p,max} = i_{m,max} = \frac{\left(V_{INmin} - 1\right) \cdot D_{max}}{L_m\, f_S} = 5.72\,A$$

Select a power MOS with a r_{DSon} such that:

$$r_{DSon} \cdot i_{m,max} \le 1V \rightarrow r_{DSon} \le 175\,m\Omega$$

Flyback regulator: design example

Step 5: determine the size of the core

$$L_m i^2_{p,max} = 1\,\mathrm{mHA}^2$$

A Kool Mμ™ distributed-gap core is suitable for our design. Going up the selector chart from 1 mHA^2 yields a 77210 core with 90μ (aggressive design). $A_L = 49\mathrm{nH/T}^2 \pm 8\%$.

Step 6: calculate primary turns N_1

$$N_1 = \sqrt{1000 \frac{L_m(\mu\mathrm{H})}{(A_L \cdot 1.08)}} = 23.8\ \mathrm{T} \qquad \text{(use 23 T)}$$

Flyback regulator: design example

Step 7: calculate magnetizing force to make sure core does not saturate

From the 77210 data sheet we see that A'_L=36.5 AT with 23 Turns. Adjust turns accordingly:

$$N_1' = N \frac{A_L}{A_L'} = 30.4\,T \qquad \text{(use 30 T)}$$

Using 30 Turns would give $L_m \sim 30\ \mu H$ in the worst condition.

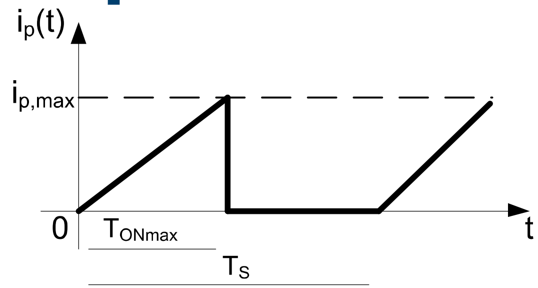
Step 8: calculate secondary turns N_2

$$N_2 = \frac{N_1'}{6} = 5\,T$$

Flyback regulator: design example

Step 9: primary wire size

$$I_{P(rms)} = \frac{i_{p,max}}{\sqrt{6}} = 2.34\ A_{rms}$$



From the wire table we see that #20 AWG is appropriate, but..

- **Skin effect**

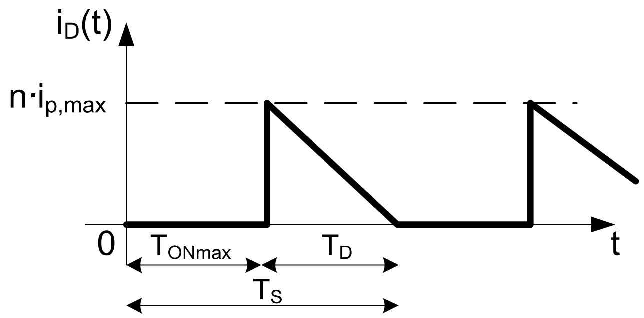
  to minimize skin effect the selected wire diameter should not exceed $2\cdot\delta$, where δ is the skin depth of copper.

- **Winding Considerations**
  - Tight coupling between the primary and secondary windings is mandatory to minimize Lleak.
  - It is best to wind the core by using a "**multi-filar**" type of winding. The bundle of wires (having the same length) can be wound on the core at one time by laying the wires "side-by-side“..
  - As a general rule, the more strands and the smaller the wire gauge, the less the leakage inductance and the tighter the coupling between windings.

Hence, use 3 strands - #25 AWG   30 Turns for primary winding.

Flyback regulator: design example

Step 10: secondary wire size



$$I_{S(rms)} = \frac{N_1}{N_2} \frac{i_{p,max}}{\sqrt{6}} \approx 14\ A_{rms}$$

Use 7 strands - #21 AWG   5 Turns for secondary winding.

Step 11: building of core

Check the winding factor, WF

$$WF = \frac{\left(N_1 \cdot n_{w1} \cdot A_{w1} + N_2 \cdot n_{w2} \cdot A_{w2}\right)}{Window\ Area} = \frac{\left(30 \cdot 3 \cdot 2 \times 10^{-3} cm^2 + 5 \cdot 7 \cdot 4.84 \times 10^{-3} cm^2\right)}{1.14\, cm^2} \cong 30\%$$

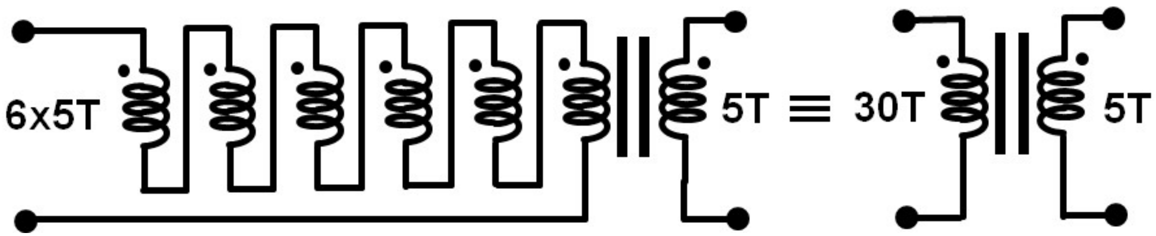
Conclusion for core design

Use 77210 core with 30T of 3 - #25AWG in parallel for the primary and 5T of 7 - #21AWG in parallel for the secondary.

Flyback regulator: design example

Step 12: windings

- We have 30T on the primary and 5T on the secondary.
- We can break the primary up into a series of sections (see figure) so that we can wind both the primary and secondary in a "*multi-filar*" fashion (single budle).
- Since 5 x 6 = 30, we need 6 series connection for the primary.
- Hence, we take 6 x 3 = 18 - #25AWG with 7 - #21AWG, combine them in a bundle and wind 5T around the core. Then connect the primary in series to make the 30 Turns.



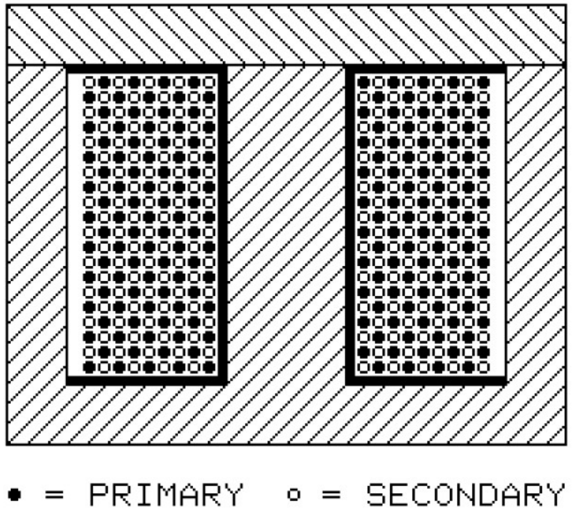
7-filar winding

5T, 6 sections in series = 30T equivalent

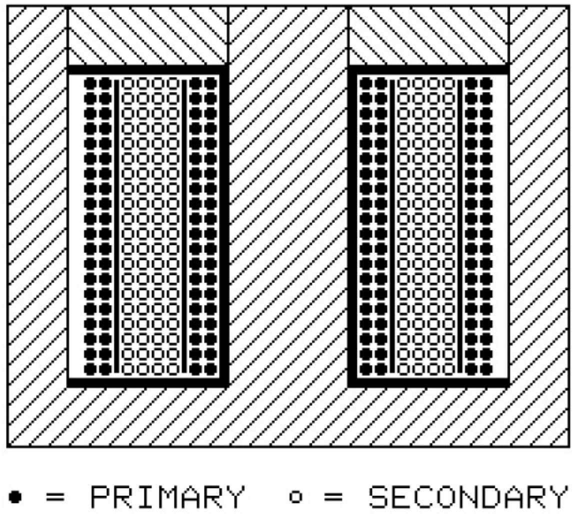
- The advantages are twofold: all wires are of equal length and the interwinding coupling is improved.
- The disadvantage is that the turns ratio must always be an integer number.

Winding Techniques:recap

- The best coupling is achieved with a one-to-one layout, meaning that the primary and secondary are wound together as a twisted-pair wire in one layer on a bobbin (multi-filar winding). This is possible, however, when the turns ratio is an integer number.



- When a fractional ratio is required, a "standard" winding technique is used, where the windings are placed seperately on the core. Interleaving the winding is a good way to go.

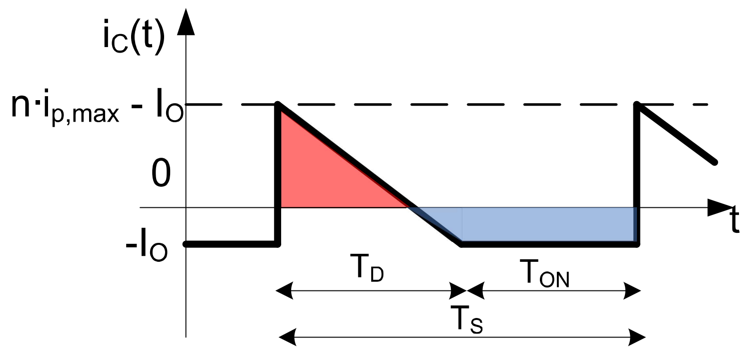


Note: for a deeper insight on transformer design see Erickson, ch.13 and Mohan, ch. 30

Flyback regulator: design example

Step 13: output capacitor

Neglecting the capacitive ripple:



$$\Delta V_O = ESR \cdot \Delta i_C = ESR \cdot \frac{N_1}{N_2} \cdot i_{p,\,max} \le 100mV \quad \Longrightarrow \quad ESR < 2.9m\Omega$$

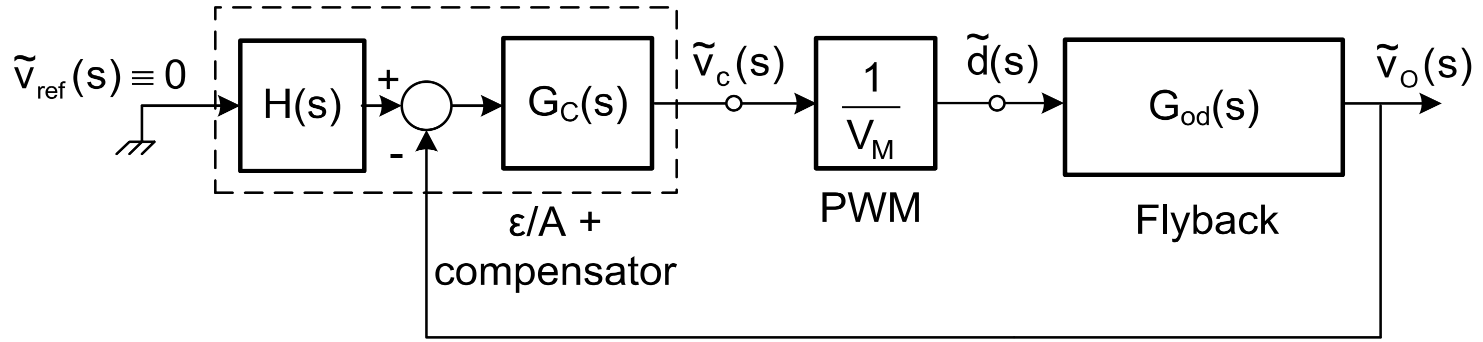
For instance, use 10 Sanyo “10SVP330M” OS-CON capacitors in parallel (C=330µF, ESR=17mΩ, rated ripple current=3.95 A_{rms}).

$$C_{TOT} = 3300\mu F,\ ESR_{TOT} = 1.7m\Omega$$

$$\boxed{\Delta v_O = \Delta v_C + \Delta v_{ESR} \approx 74mV}$$

Flyback regulator: design example

Step 14: closing the feedback loop (voltage mode control)



In DCM:

$$G_{od}(s) = \frac{\tilde{v}_O(s)}{\tilde{d}(s)} = G_{od}(0)\frac{1+\dfrac{s}{2\pi f_z}}{1+\dfrac{s}{2\pi f_p}}$$

$$f_z = \frac{1}{2\pi \cdot ESR \cdot C} \cong 28KHz$$

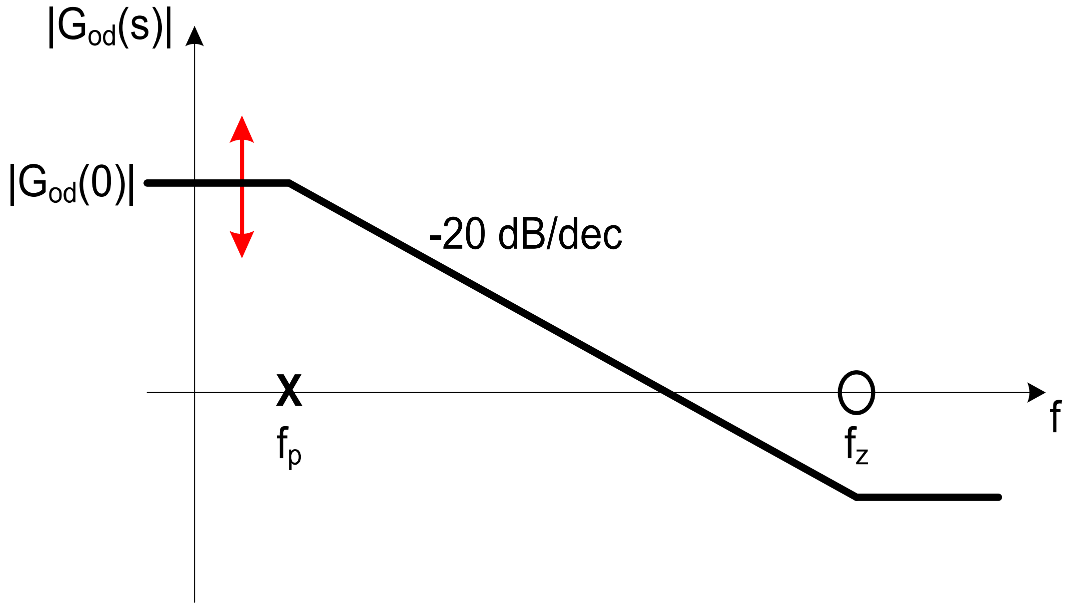
$$G_{od}(0) = \frac{\tilde{v}_O(0)}{\tilde{d}(0)} = V_{IN}\sqrt{\frac{R}{2L_m \cdot f_s}}$$

$$\boxed{f_p = \frac{2}{2\pi \cdot R \cdot C} = \frac{1}{\pi \cdot R \cdot C}}$$

Flyback regulator: design example

Note

- $G_{od}(0)$ depends on V_{IN}
- the pole frequency, f_p, is fixed:
  - ➢ with I_O = 8 A we have R = 0.625 Ω, giving $f_p \approx 154$ Hz



Flyback regulator: design example

Control-to-output transfer function

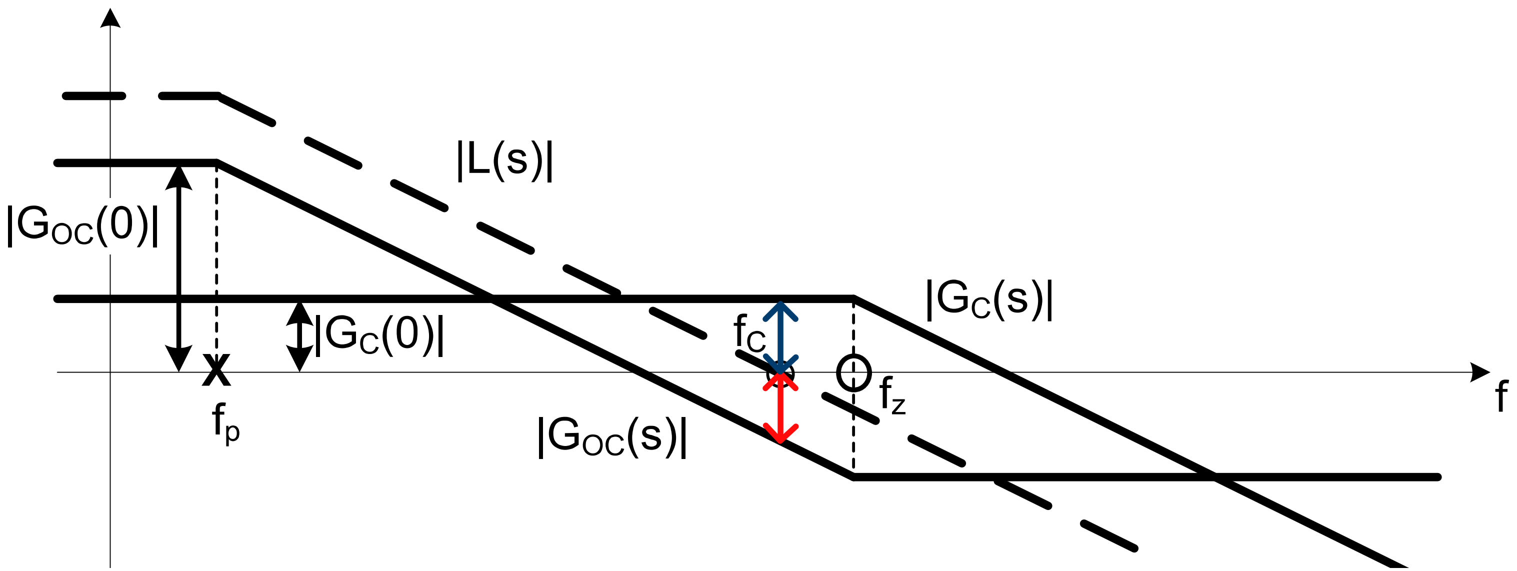
$$G_{OC}(s)=\frac{\tilde{v}_O(s)}{\tilde{v}_C(s)}=G_{od}(s)\cdot G_M(s)=\frac{G_{od}(0)}{V_M}\frac{1+\frac{s}{2\pi f_z}}{1+\frac{s}{2\pi f_p}}$$

For instance, V_M=2.5 V for a UC1524A controller. Therefore:

$$G_{OC}(0)=\frac{\tilde{v}_O(0)}{\tilde{v}_C(0)}=\frac{V_{IN}}{V_M}\sqrt{\frac{R}{2L_m\cdot f_s}}\cong 0.129V_{IN}$$

- $\boxed{G_{OC}(0)\big|_{max}=9.29 \text{ for } V_{IN}=V_{INmax}}$
- $G_{OC}(0)\big|_{min}=4.64 \text{ for } V_{IN}=V_{INmin}$

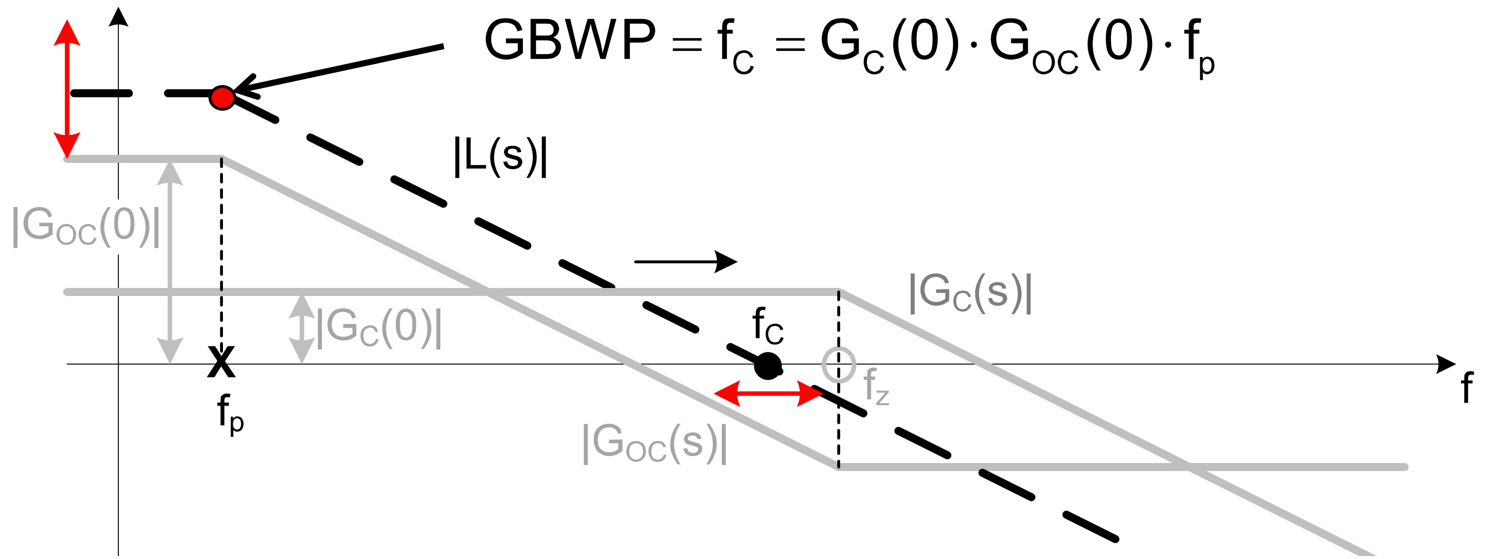
Flyback regulator: design example



Simple first-order low pass compensation network

- compensation pole used to cancel ESR zero
- set the crossover frequency (0 dB loop gain) at the desired value ($f_c = f_s/5 = 20$ kHz) by adjusting $G_C(0)$

Flyback regulator: design example



- Worst condition for for $G_{OC}(0) = G_{OC}(0)\big|_{max} = 9.29$

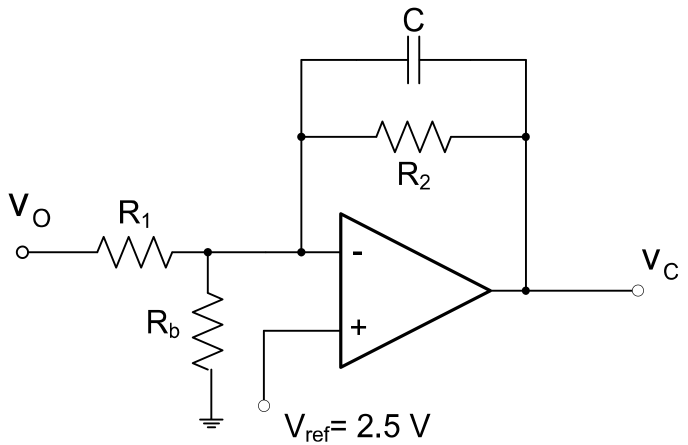
- Using GBWP

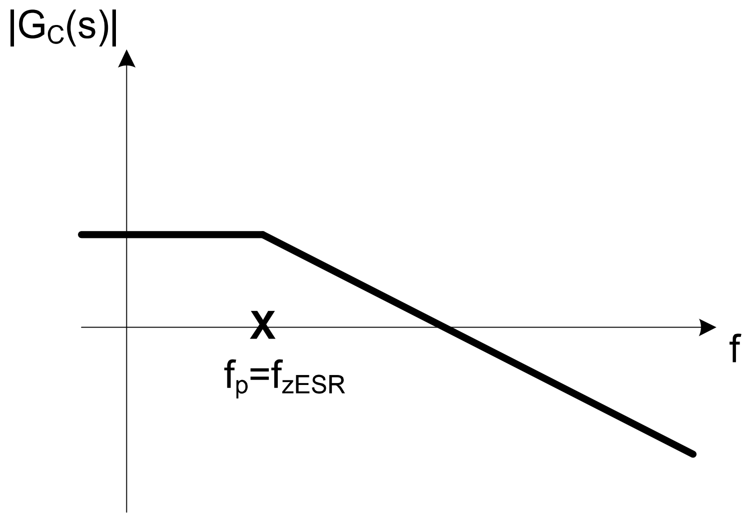
$$\boxed{f_{C_{MAX}} = G_C(0) \cdot G_{OC}(0)\big|_{max} \le \frac{f_{SW}}{5} = 20\text{kHz}}$$

$$\left\{ \begin{aligned} & G_C(0) \le 13.96 = 23\,\text{dB} \\ & G_{OC}(0)\big|_{max} = 9.3 = 19.35\,\text{dB} \\ & \big|L(0)\big|_{max} = 42.35\,\text{dB} \end{aligned} \right.$$

Flyback regulator: design example

- Compensation network



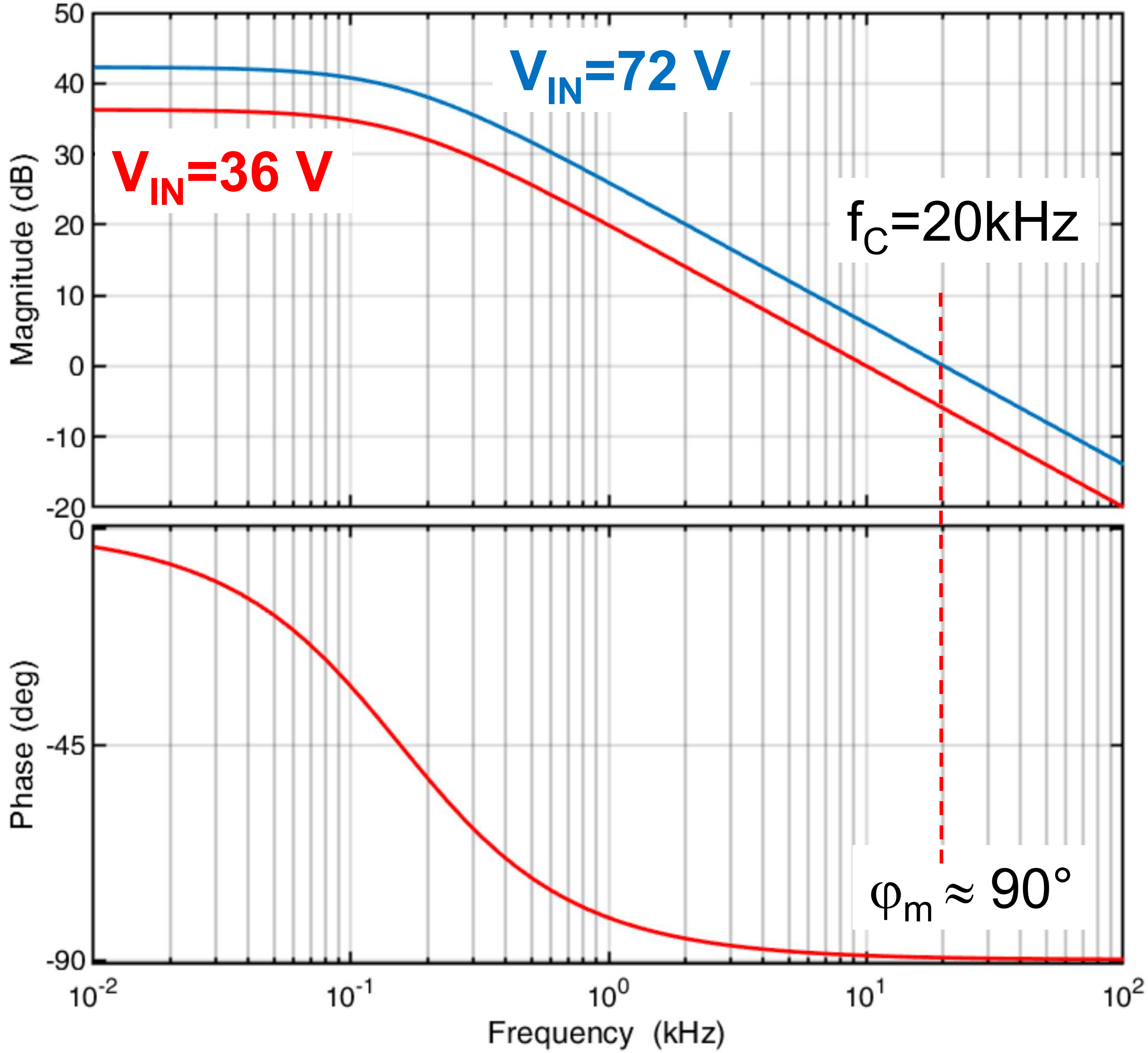


$$G_C(0) = \frac{R_2}{R_1} \qquad f_{zESR} = \frac{1}{2\pi R_2 C}$$

- Use for instance:
  - ➢ $R_1 = 2.8\ k\Omega$
  - ➢ $R_b = 2.8\ k\Omega$
  - ➢ $R_2 = 39\ k\Omega$
  - ➢ $C = 150\ pF$

May want to consider a second capacitor in series with R_2

L(s): Bode diagram



Flyback regulator: design example

- Efficiency check

$$\eta = \frac{P_O}{P_O + \sum P_{loss}} \quad \text{with} \; P_D = I_{O,n\,max} \cdot V_D \approx 8\,W \;,\; P_{TRANSF} \approx 1.7\;W,\; P_{sw} = 0.9W$$

resulting in $\eta = \frac{P_O}{P_{IN}} = 0.79$! No need for a 2nd iteration.

Note: power dissipation mainly comes from the diode. Solutions?

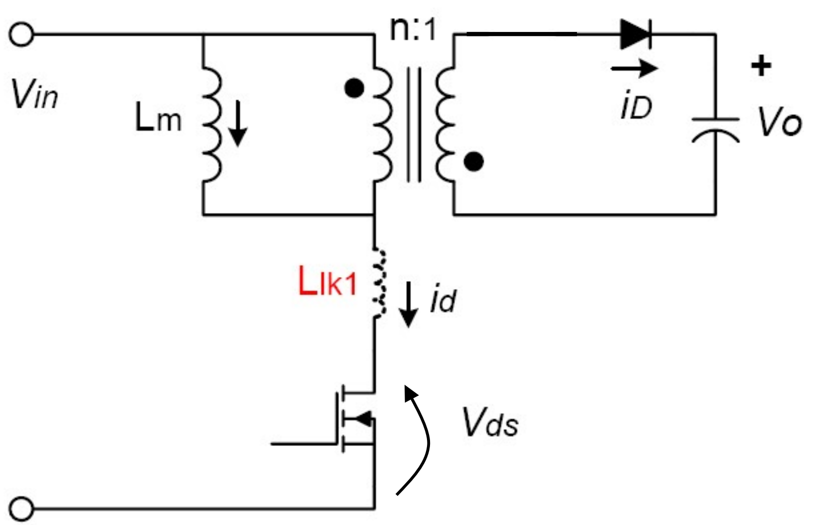
- Minimum duty-cycle check

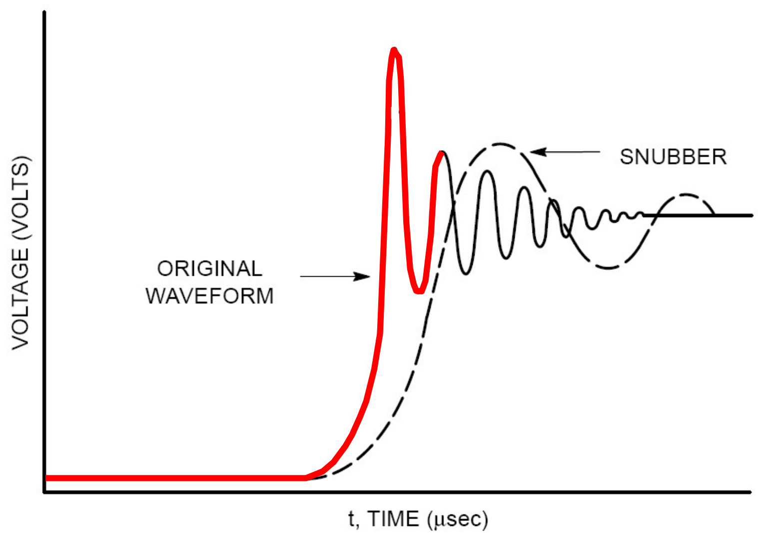
From $\frac{V_O}{V_{IN}} = D\sqrt{\frac{\eta \cdot R}{2L_m f_S}}$, we get $\boxed{D_{min} = \frac{V_O}{V_{INmax}}\sqrt{\frac{2L_m f_S}{\eta \cdot R}} \approx 24\%}$

Flyback regulator: leakage inductance

- Leakage inductance overvoltage

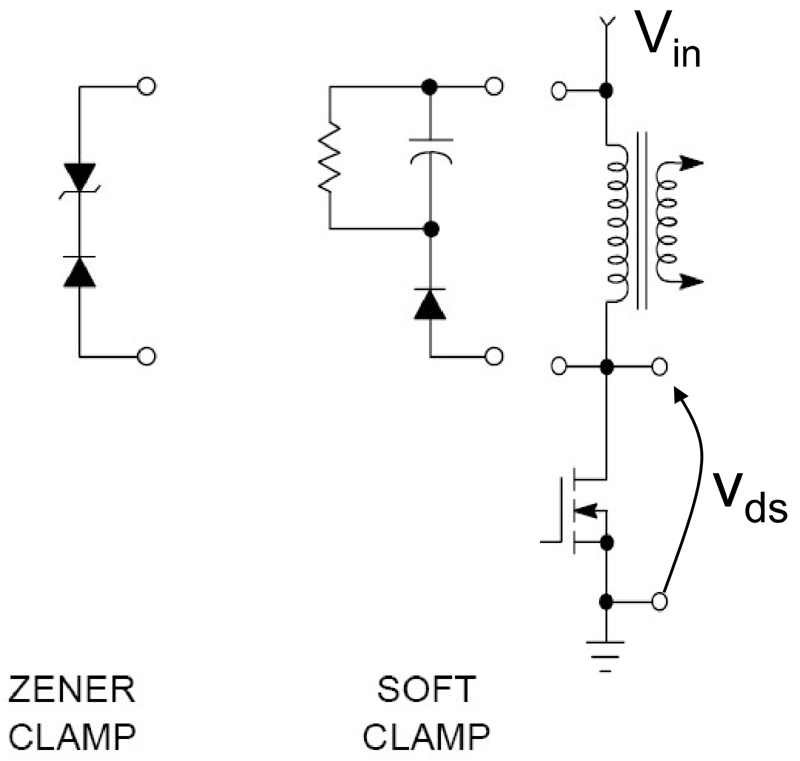
When the power MOSFET is turned off, there is a high voltage spike on the drain due to the transformer leakage inductance. This excessive voltage on the MOSFET may lead to an avalanche breakdown and eventually failure of power MOSFET. Therefore, it is necessary to use an additional network to clamp the voltage.

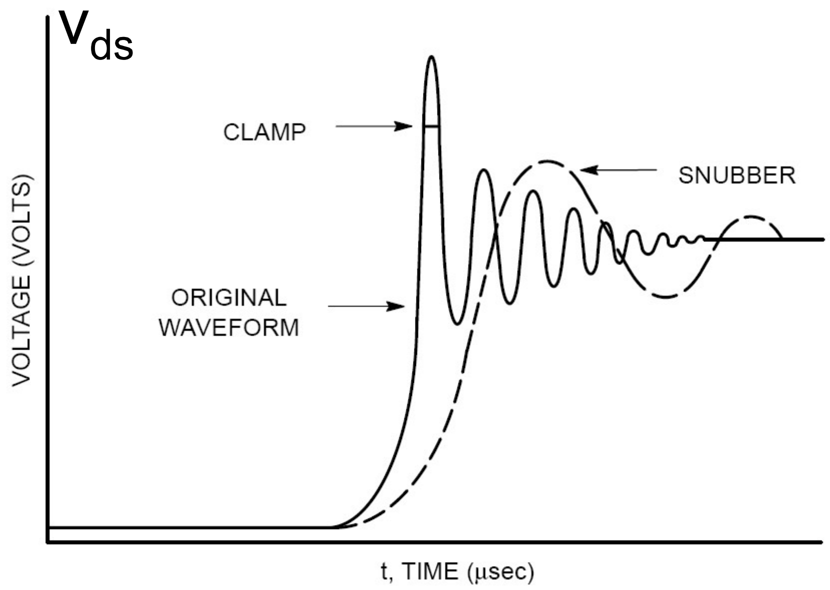




Flyback regulator: leakage inductance

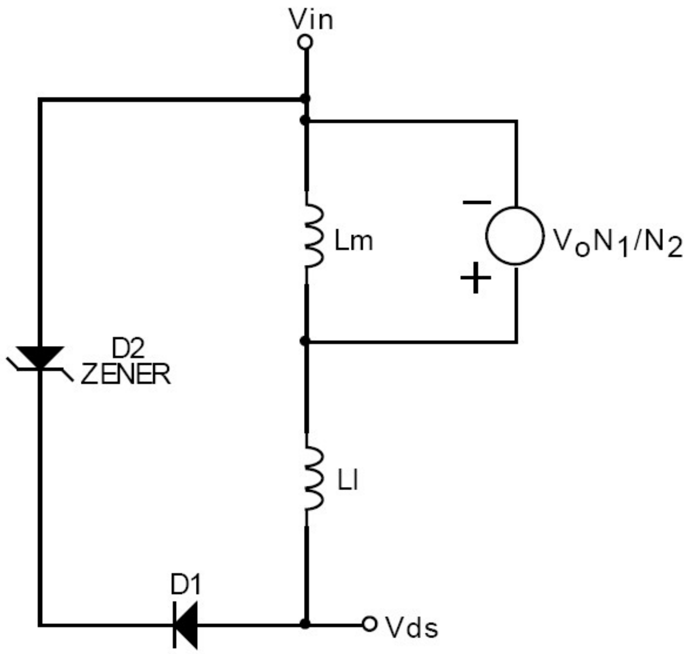
- Clamp networks

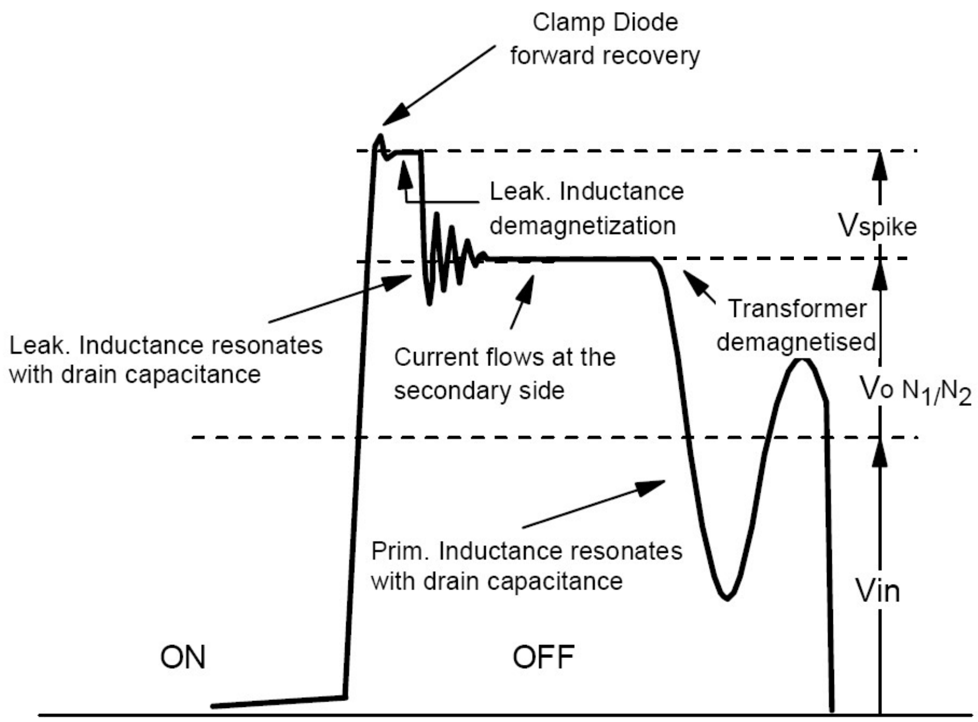




Flyback regulator: leakage inductance

- Zener clamp





Flyback regulator: leakage inductance