PASSIVE LOSSLESS SNUBBERS FOR HIGH FREQUENCY PWM CONVERTERS

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OUTLINE

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References

Chapter 1

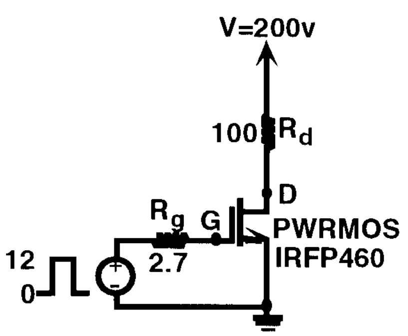
INTRODUCTION

SEMINAR OBJECTIVES

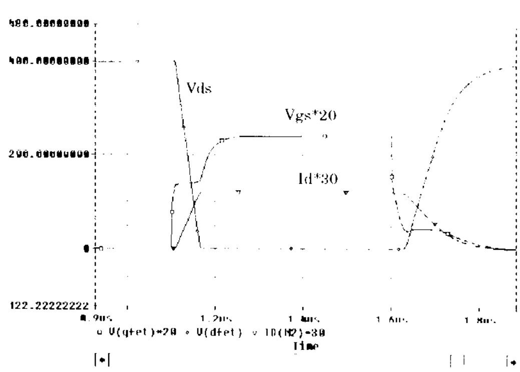
- To demonstrate the use and benefits of passive lossless snubbers in modern high frequency power electronics design.

  - Relevant topologies
  - Soft switching
  - Parasitic effects
  - Limitation - pros & cons

HARD VERSUS SOFT SWITCHING

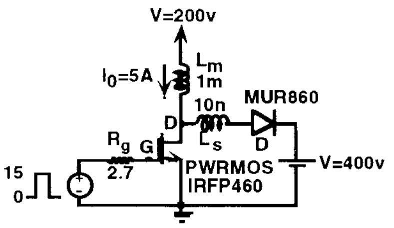


MOSFET switching stage

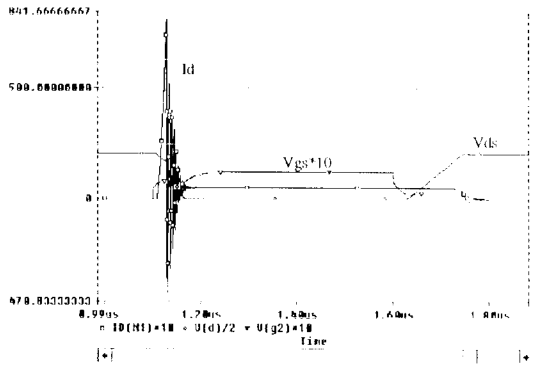


Switching waveforms (simulation)

- Switching losses are proportional to switching frequency
- Modern simulators include switching behavior



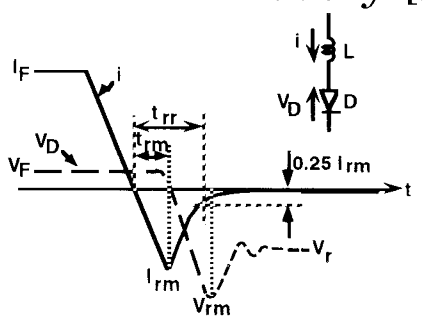
Hard switched boost stage



Basic waveforms (simulation)

- ☛ Diode reverse current and parasitic oscillations
- ☛ Natural 'softening' at 'turn off'
  - 'Turn On' more troublesome in MOSFETS
  - 'Turn On' & 'Turn off' problematic for IGBT -> ZCS

Diode Reverse Recovery [25, 39]



- Low voltage drop on diode in reverse recovery period
- $\dfrac{di}{dt} = \dfrac{V}{L}$
- High reverse currents
- High reverse voltage
- Parasitic oscillations

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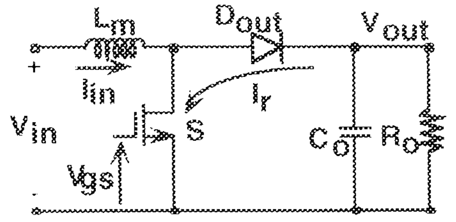
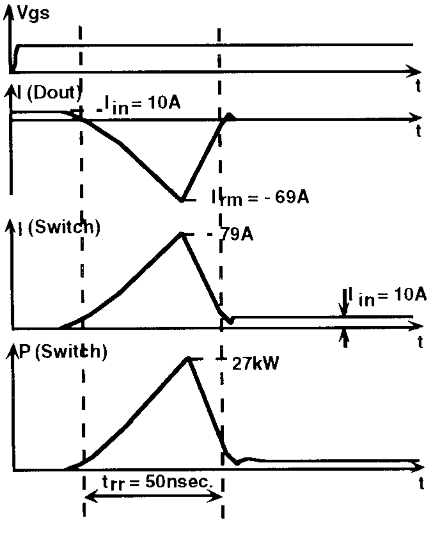
The reverse recovery problem in Boost topology



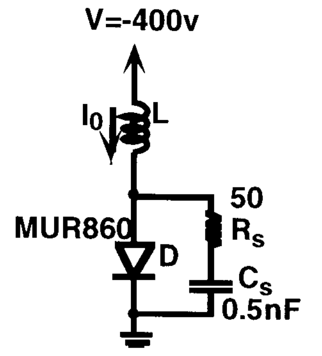
- Short circuit against V_{out}
- Occurs in all topologies

The reverse recovery problem in Boost topology



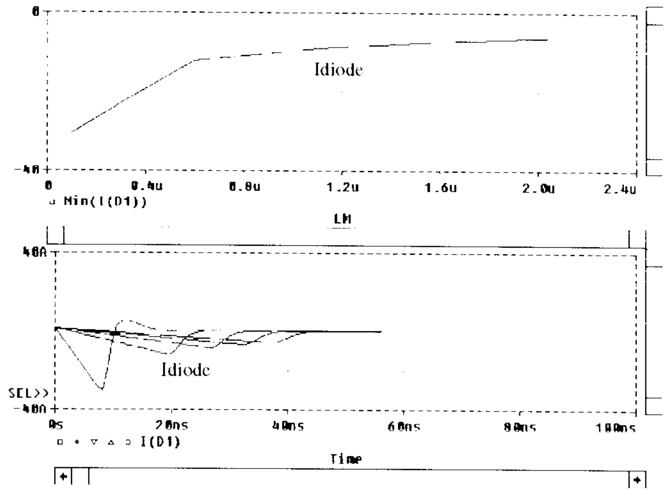
- High reverse currents
- High losses
- EMI emission

Dependence on L



Diode switching circuit - reverse recovery

Dependence on L - Simulation



Diode reverse recovery switching waveforms

- Very high reverse currents
- A function of series inductance
- Detrimental effect in isolated and non isolated converters
- Becomes very important in high switching frequency converters

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Trapped Energy as L becomes Larger

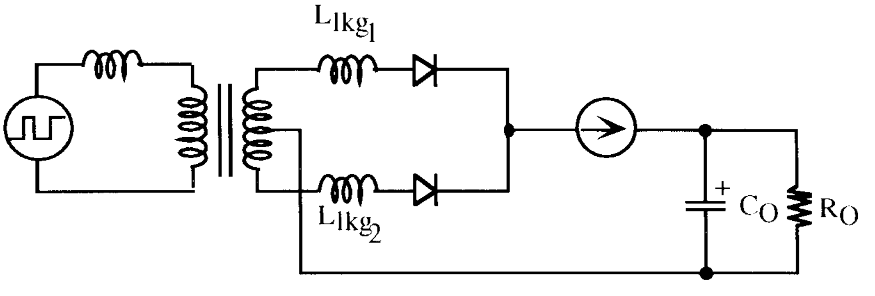
Assume: t_{rm} constant

Peak reverse current : $$I_{rm} = \frac{V}{L} t_{rm}$$

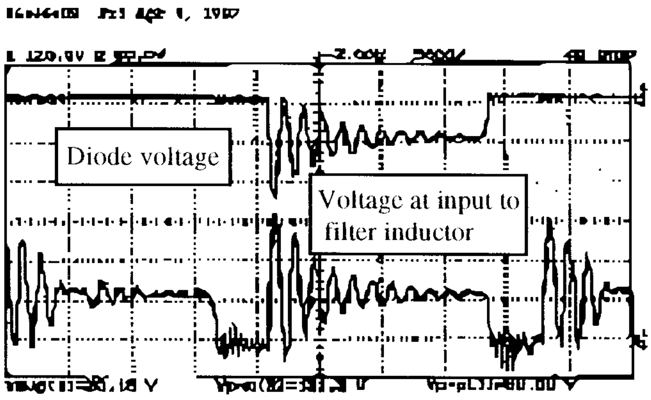
Trapped energy $$E = \tfrac{1}{2}(I_{rm})^2 L = \frac{(V t_{rm})^2}{2L}$$

- Trapped energy decreases as L becomes larger
- Adding inductance ---> Advantageous

Center Tap Rectifier



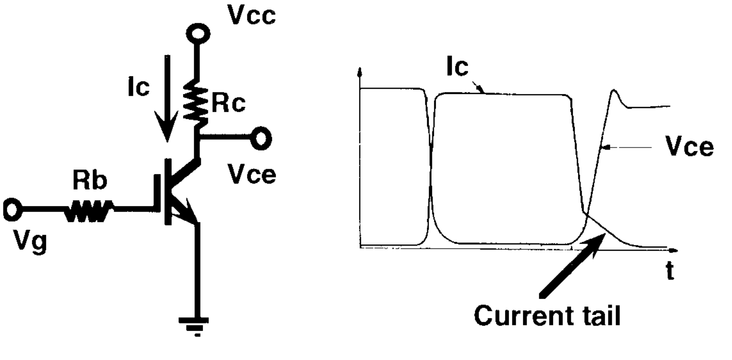
Leakage inductance at secondary



Experimental waveforms (no diode snubber)

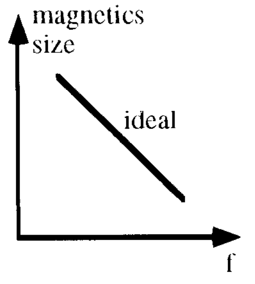
- Greatly influenced by transformer leakage
- Forces the use of high voltage diodes

IGBT BEHAVIOR UNDER HARD SWITCHING
[5, 16, 20, 21, 38]

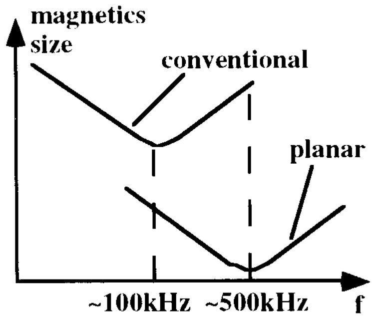


- Current tail due to stored minority carriers
- Switching losses increase linearly with switching frequency
- Switching frequency limit is at about 25 kHz

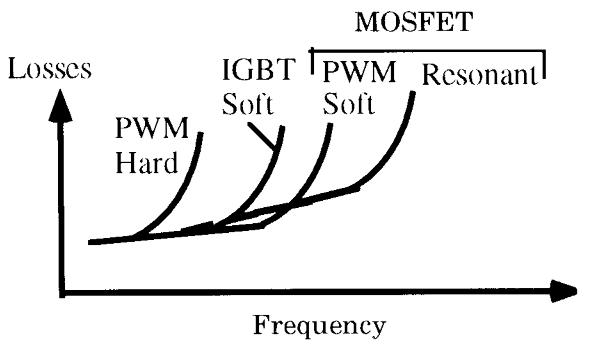
Why Soft Switching ?



Ideal Size/Switching frequency relationships



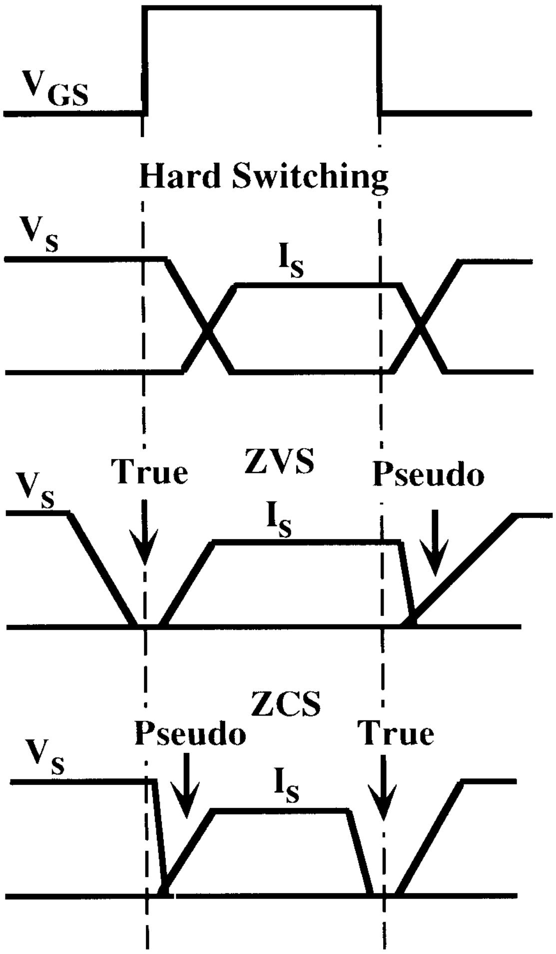
Practical size realization



Losses

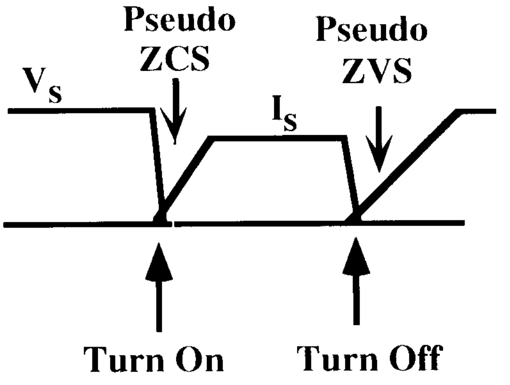
- Conflicting constraints
- Soft switching helps to reduce EMI emission ???

SOFT SWITCHING TERMINOLOGY



- 'True' soft switching calls for external active circuitry
- 'Pseudo' soft switching by lossless snubbers

Soft Switching by Passive Lossless Snubbers



- For MOSFETs -> OK
- For IGBTs -> ZCS at 'turn off' might be more desirable (current tail)
- This can not be accomplished with a Passive Snubber unless operating in constant 'off' time (quasi-resonant)

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SIMULATION TOOLS

✔ Modern device models account for switching behavior
- ☛ Old Models DO NOT account for switching behavior
- ☛ Difficult to account for PCB parasitics
- ☛ Simulation of switching losses is still unreliable

☆ Cycle-by-Cycle simulation
- ✔ Can faithfully describe the basic switching phenomena
- ✔ Very useful tool to examine snubber operation

☆ Average Simulation [3]
- ✔ Can account for snubber effect on dynamics
- ✔ Can be used to design the feedback loop

MODEL MUR860 D (IS=783U RS=30M N=4.82 BV=600 IBV=10U
+ CJO=330P VJ=.75 M=.333 TT=79.2N)
* Motorola 600 Volt 8 Amp 55M us Si Diode 02-24-1994

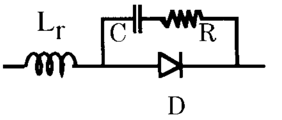
.MODEL DN5406 D (IS=2.68P RS=7.31M N=1.17 BV=900 IBV=10U
+ CJO=124P VJ=.6 M=.333 TT=14.4U)
* Motorola 600 Volt 3 Amp 15 us Si Diode 11-23-1990

- Parameters can be changed to test various aspects, e.g.TT for reverse recovery

```
.SUBCKT IRF150   10 20 30
*   TERMINALS:  D  G  S
M1   1  2  3  3  DMOS L=1U W=1U
RG  20  2  5.35
RD  10  1  4.3M
RDS  1  3  635K
CGD  4  1  2.75N
RCG  4  1  10MEG
MCG  4  5  2  2  SW L=1U W=1U
ECG  5  2  2  1  1
DGD  2  6  DCGD
MDG  6  7  1  1  SW L=1U W=1U
EDG  7  1  1  2  1
DDS  3  1  DSUB
LS  30  3  7.5N
.MODEL DMOS NMOS (LEVEL=3 VMAX=1.6MEG THETA=265.6M
VTO=3.3
+ KP=9 RS=8.12M IS=2.01P CGSO=2.65M)
.MODEL SW NMOS (LEVEL=3 VTO=0 KP=.45)
.MODEL DCGD D (CJO=2.75N M=.5 VJ=.41)
.MODEL DSUB D (IS=2.01P RS=5.20M VJ=.8 M=.4 CJO=2.60N
TT=720N)
.ENDS
*  IR 100 Volt  28 Amp  45M Ohm  N-Channel Power MOSFET  11-20-
1990
```

Chapter 2

PASSIVE LOSSLESS SNUBBERS PERSPECTIVE



RC SNUBBERS - Diode RC snubber

LOSSES $P_{d(min)} = \{\frac{(2V_o)^2 C}{2}\} f_s$; $P_{d(lkg)} = \{\frac{(I_{pkr})^2 L_{lkg}}{2}\} f_s$

$P_{d(min)}$ = minimum losses

V_0 = output voltage

C = snubber capacitor

f_s = switching frequency

I_{pkr} = peak reverse current

L_{lkg}= leakage inductance

Resistor dissipation may reach 10's of Watts

PASSIVE LOSSLESS SNUBBER APPROACHES

1. Snubbing in multiple switch configuration (e.g. half bridge)
2. Auxiliary (dual) switch snubbers [4, 15, 33, 37]
3. Snubbers in single switch configuration (references given below)

No. # 1: Most effective

- Few extra components => economical

No. # 2: Effective but costly

- Calls for extra switches and drives
- Diminishing return

No. # 3: Good compromise

- Only passive elements needed
- Proven to improve efficiency
- Potentially lowers EMI emission

This seminar will concentrate on single switch and diode snubbers for PWM converters

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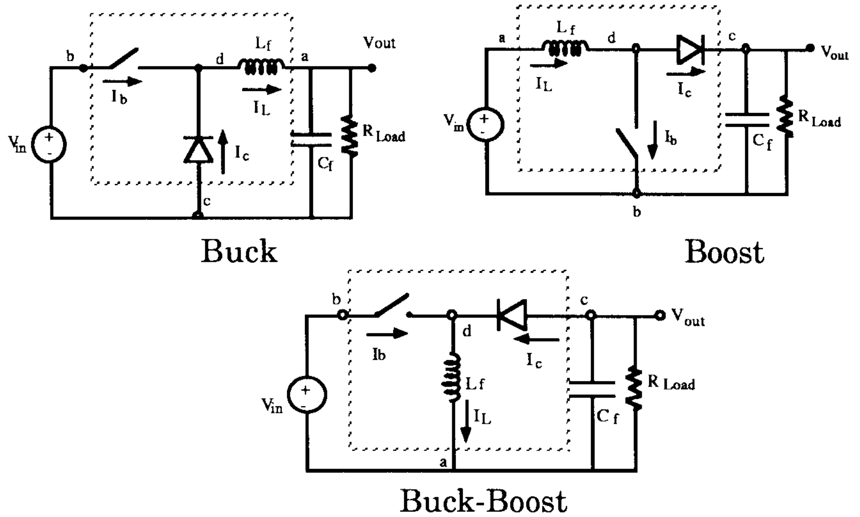
Snubbers Evolution

- SCR (1): RC snubbers - high losses
- SCR (2): Lossless Snubbers - help to turn off devices
- PWM (1): RC Snubbers - high losses
- PWM (2): Quasi Resonant - high stresses and conduction losses
- PWM (3): Auxiliary Switch - complex
- PWM (4): Lossless Snubbers -optimal (as of now)

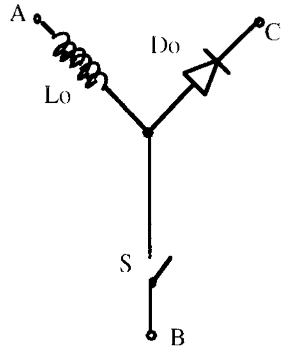
☞ Passive Lossless Snubbers are comparable in performance to the Dual Switch approaches but are less expensive !

The Switched Inductor (SIM) Model

BASIC SWITCHING CELL OF COMMON PWM CONVERTERS



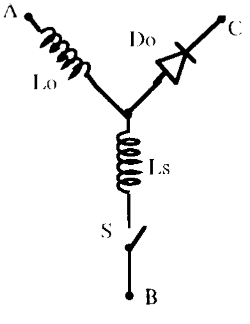
BASIC SWITCHING CELL OF COMMON PWM CONVERTERS



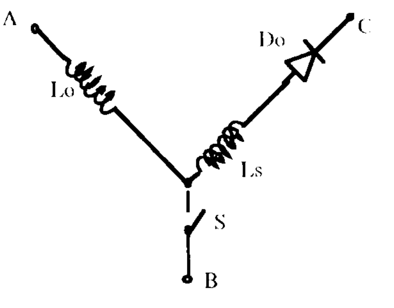
Topology	V_{in}	V_o	Topology	V_{AB}	V_{CB}
Buck	V_{CB}	V_{CA}	Buck	$V_{in}-V_o$	V_{in}
Boost	V_{AB}	V_{CB}	Boost	V_{in}	V_o
Buck-Boost	V_{AB}	$-V_{CA}$	Buck-Boost	V_{in}	$V_{in}+V_o$

Fundamental Principles

1. Controlling $\frac{dI}{dt}$ at 'Turn On' -> Pseudo ZCS



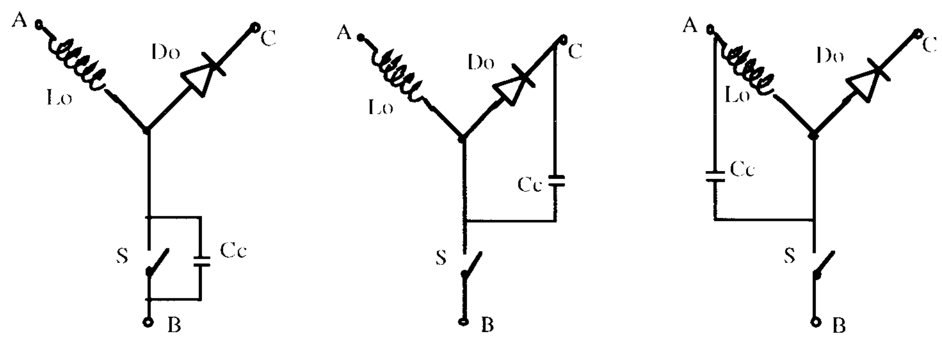
S - type
Turn-On snubber



D - type
Turn-On snubber

Fundamental Principles

2. Controlling $\frac{dV}{dt}$ at 'Turn Off' -> Pseudo ZVS



S - type Turn-Off snubber | D - type Turn-Off snubber | L - type Turn-Off snubber

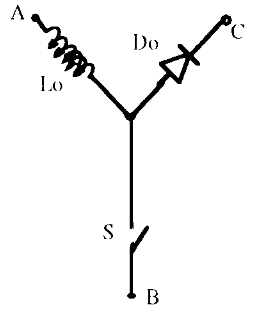
The major objective: to circulate the trapped energy

- In a lossless manner
- Without increasing the switch and diode stresses
- As quickly as possible (D_{on} & D_{off} limitations)
- Without generating new parasitic effects (of extra components)
- Inexpensive to implement

Use the followings as check points for comparison

I_{pk}(switch)
V_{max}(switch)
V_{max}(diode)

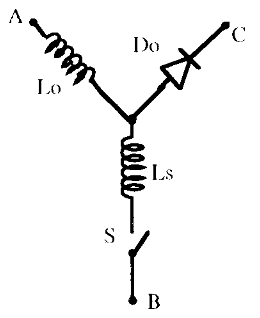
General Observation



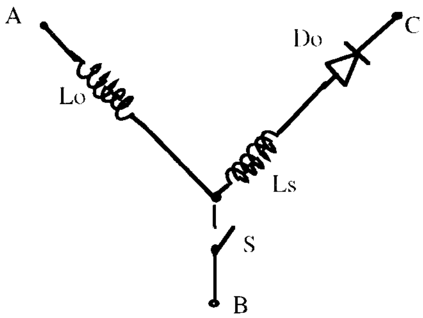
Snubber solution is independent of (PWM) topology if confined to the 'A-B-C' domain

Practical Aspects (1)

Snubber inductor rms current:



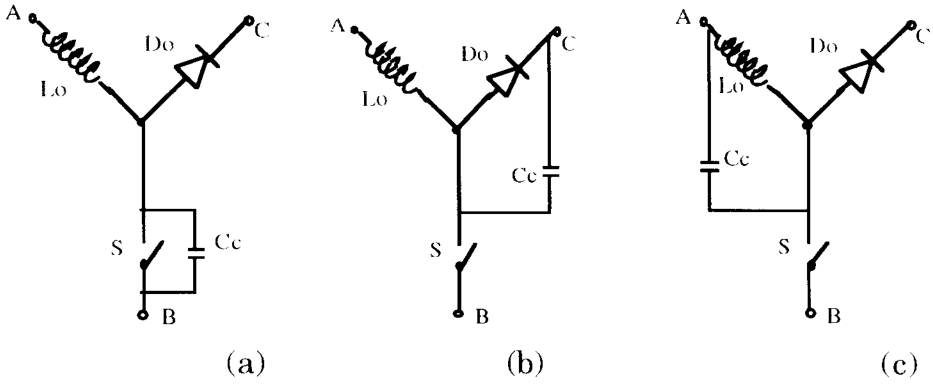
- Snubber inductor carries switch plus reverse recovery currents



- Snubber inductor carries diode plus reverse recovery currents

Practical Aspects (2)

Stirring capacitor's ripple current



(a) Ripple stirred to 'ground' - in Boost topology
(b) Ripple stirred to output - in Boost topology
(c) Ripple stirred to input - in Boost topology

Chapter 3

BASICS OF RESONANT NETWORKS

Reset of Resonant Elements

At steady-state:

- Volt-Sec of resonant inductor over switching cycle must be zero

$$\int_{T_s} V_{L_r} dt = 0$$

- Ampere- Sec of resonant capacitor over switching cycle must be zero

$$\int_{T_s} I_{C_r} dt = 0$$

- Sufficient time must be available for reset
  - This will cause restriction on duty cycle min & max

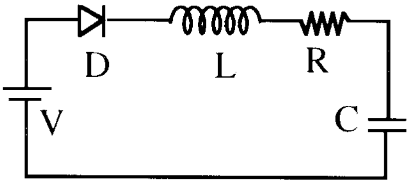
Resonant Networks

-the vehicle of Snubbing and Energy Circulation

- Energy circulation in passive lossless snubbers is made possible by lossless energy exchange of reactive elements (i.e. C, L)

- When the exchange is between L & C one deals with resonant phenomena

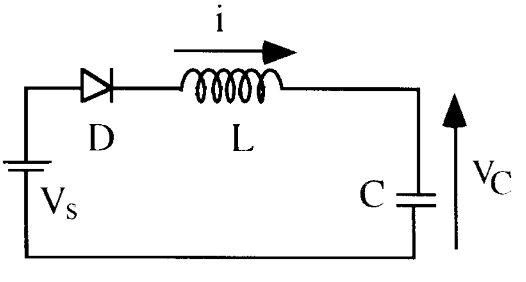
Basic Resonant Network Parameters



- Typical equivalent circuit of a lossless snubber

➜ Series resonance
➜ R is normally small (to make the snubber "lossless")
➜ May or may not include diode
➜ Peak current equal or higher than main currents
➜ Resonant frequency need to be shorter than switching frequency

Basic Resonant Network Parameters

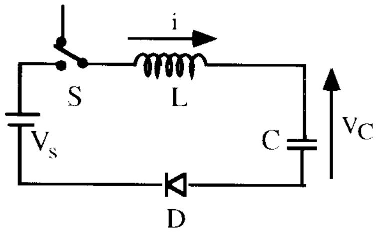


➜ Resonant frequency: $f_r = \frac{1}{2\pi\sqrt{LC}}$; $\omega_r = \frac{1}{\sqrt{LC}}$; $T_r = \frac{2\pi}{\omega_r}$

➜ Characteristic impedance : $Z_r = \sqrt{\frac{L}{C}}$

Ideal LC-network with ideal diode fed by a voltage source V_s

$i(0)= I_o \quad v_C(0)=V_{Co}$



$$i=\frac{V_s-V_{Co}}{Z_r}\sin(\omega_r t)+I_o\cos(\omega_r t)$$

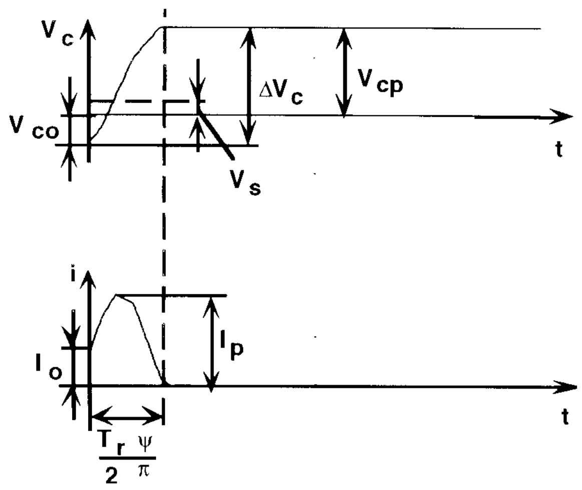
$$v_C=(V_s-V_{Co})[1-\cos(\omega_r t)]+I_oZ_r\sin(\omega_r t)+V_{Co}$$

$$V_{Cp}=(V_s-V_{Co})[1-\cos\psi]+I_oZ_r\sin\psi+V_{Co}$$

$$\psi=\tan^{-1}\left(-\frac{I_oZ_r}{V_s-V_{Co}}\right)+\pi \qquad \psi<\pi$$

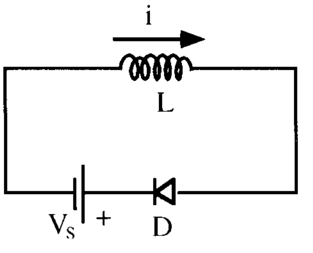
If $V_{Co} = 0$ and $I_o = 0 \quad \psi=\pi$ and $V_{Cp} = 2\ V_s$

If $V_{Co} < 0$ and $I_o = 0 \quad \psi=\pi$ and $V_{Cp} > 2\ V_s$



SOME TRIVIAL CASES

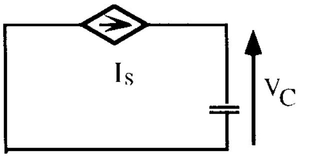
- Linear inductor discharge



$$\frac{dI_L}{dt} = \frac{V_S}{L}; \qquad i_L = I_L(0) - \frac{V_S}{L} t$$

✌ Large capacitor acts as V_S

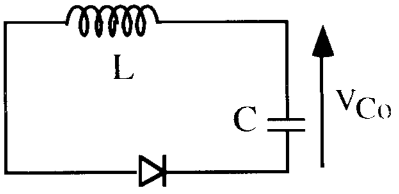
- Linear capacitor discharge



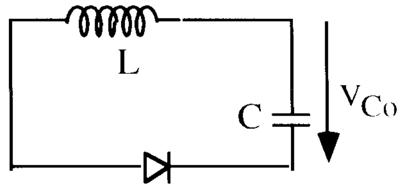
$$\frac{dV_C}{dt} = \frac{I_S}{L}; \quad v_C = V_C(0) + \frac{I_S}{C} t$$

✌ Large inductor acts as I_S

- Capacitor voltage reversal $i(0)=0$ $v_C(0)=V_{Co}$



Beginning

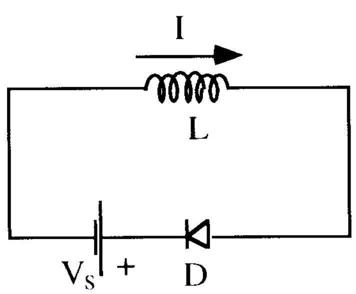


End

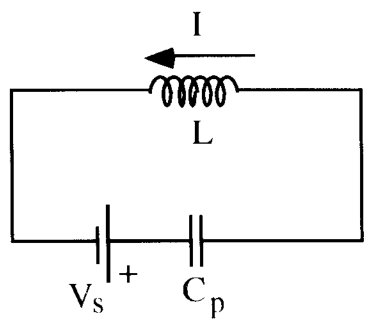
Practical Aspects (3)

Practical Snubber Components

- Reverse recovery of snubber diodes



Before reset



Diode snapped

👁 Parasitic oscillations

✔ Remedy -> Saturable Reactor

Practical Aspects (4)

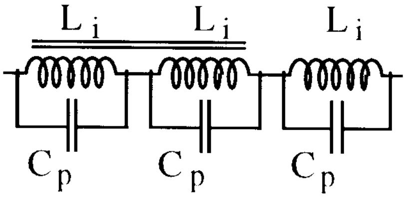
Practical Snubber Components

- ESR of resonant Capacitor

High losses, high temperature

✔ Remedy -> Low ESR capacitors:
Polypropylene
Mica (the best !!)

- Interwinding capacitance of resonant inductor



Parasitic oscillations , high losses, EMI emission

✔ Remedy -> Careful design, toroidal configuration, good coupling between windings

- Printed circuit layout

Parasitic oscillations , high losses, EMI emission

Remedy -> Careful design

Resonant Inductor Design

Typical characteristics

- Low inductance ; Typical range 3μH - 10μH
- High current ; Typical range 1A -30A

Some basic relationships:

$$B_{\max} = \frac{L\, I_{pk}}{n A_e}; \qquad L = \frac{n^2 \mu_o \mu_r A_e}{l_e};$$

$$\Delta B = \frac{\int V_L dt}{n A_e}$$

L - inductance (H)

I_{pk} - peak inductor current (A)

V_L - voltage across inductor (V)

t - time (Sec)

n - number of turns

B_{max} - limit of magnetic flux density (T)

A_e - effective core area (m^2)

l_e - effective magnetic length (m)

μ_o - permeability ($1.25\ 10^{-6}$ H/m)

μ_r - relative permeability

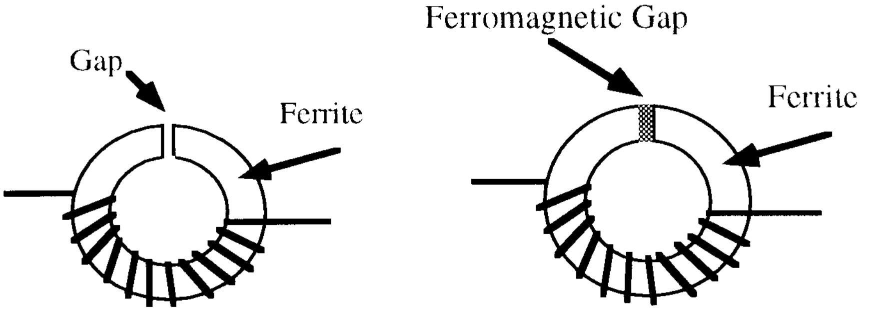
From above

$$\frac{I_{pk}}{L} = \frac{1}{L}\frac{B_{max}\, l_e}{n\,\mu_o\,\mu_r}$$

For high $\frac{I_{pk}}{L}$ ratios:

- Long l_e
- Small number of turns n
- Low relative permeability μ_r
- Winding window will no be full

Typical Construction



Toroid based design

Lower EMI emission

Suggested Design Procedure

1. Choose ΔB_{max} based of acceptable losses (mW/gr) from ferrite data (losses as a function of ΔB and frequency). Use f_r as an indicator but take into account the short period of snubbing by dividing the data sheet losses by (f_r/f_s)
2. Estimate {V Sec} across resonant inductor $\int V_L dt$ or by an approximation e.g. $V_L * t_{rr}$
3. Calculate nA_e from the relationship:

$$nA_e = \frac{\int V_L dt}{\Delta B_{\max}}$$

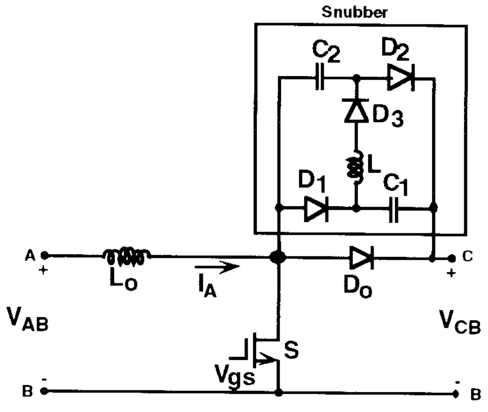
4. Select wire cross section according to rms current
5. Based on calculated nA_e and wire size, choose a core for a single layer winding
6. Gap the core for required L value

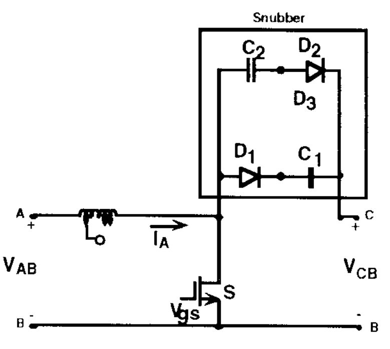
Chapter 4

SWITCH TURN-OFF LOSSLESS SNUBBERS

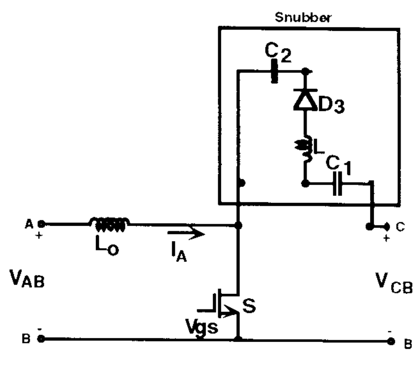
SWITCH 'TURN OFF' LOSSLESS SNUBBER (SNB1)

The 'One Way' Capacitor : Version 1 [35]





Snubbing

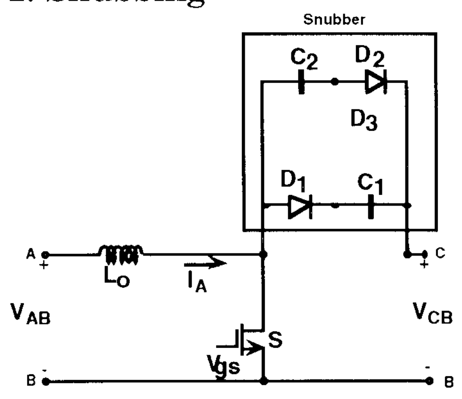


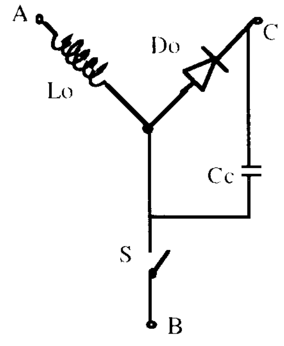
Reset

- Snubber capacitor C_1+C_2
- Resonant reset (C_1+C_2, L)

General Observations (1)

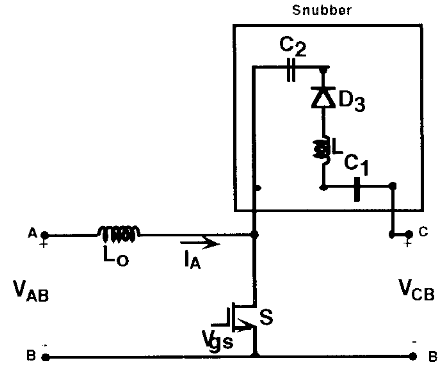
1. Snubbing

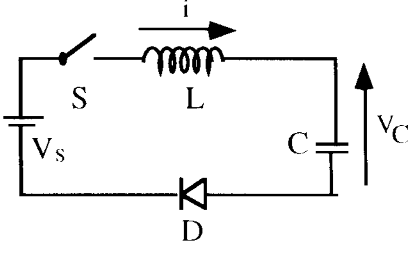




- Prior to 'turn-off', C_1 & C_2 must be charged to V_{CB}

2. Reset

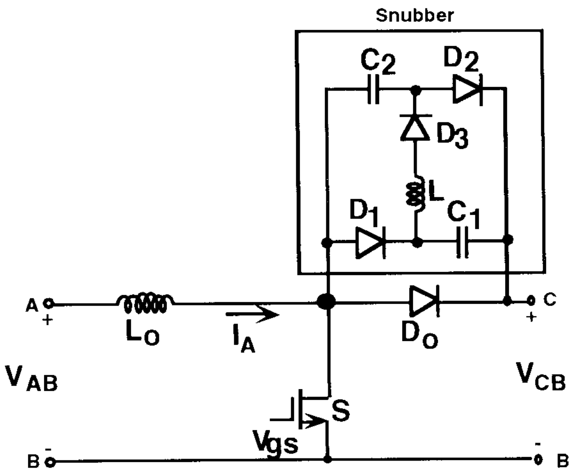


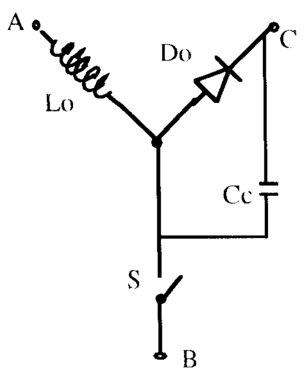


$V_C = 2\ V_{CB}$

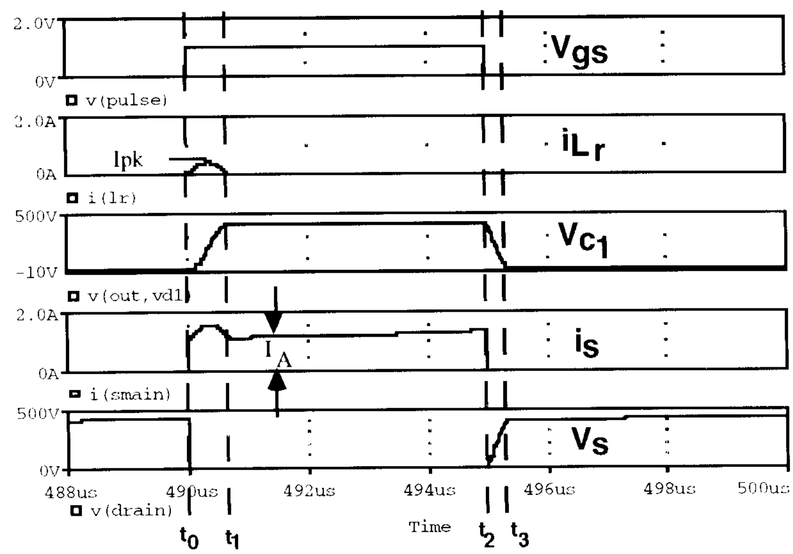
if $C_1 = C_2$ -> $V_{C1} = V_{C2} = V_{CB}$

General Observations (2)





- Topology independent



Waveforms of SNB1 (simulation)

Interval t_o-t_1

Assuming: $C_1=C_2=C$

$$t_1-t_o=\frac{T_r}{2}=\pi\sqrt{\frac{LC}{2}} \;;$$

$$i_L=i_{D3}=I_{pk}\sin(2\pi\frac{t}{T_r})$$

$$I_{pk}=\frac{V_{CB}}{\sqrt{\frac{2L}{C}}} \quad ; \quad i_S=I_A+i_L$$

$$v_{C1}=v_{C2}=0.5V_{CB}[1-\cos(2\pi\frac{t}{T_r})]$$

Interval t_1-t_2 $\quad i_S=I_A \;; \quad i_{Do}=i_{D1}=i_{D2}=i_{D3}=0$

Interval t_2-t_3 $\quad v_{C1}=v_{C2}=V_{CB}-\frac{I_A}{2C}(t-t_2) \;;$

$v_{C1}(t_3)=v_{C2}(t_3)=0$

from which: $t_3-t_2=\frac{2CV_{CB}}{I_A}$ and hence $(\frac{dv_S}{dt})_{t_2\text{-}t_3}=\frac{I_A}{2C}$

Interval t_3-(t_o+T_s) $\quad i_{Do}=I_A \;; i_S=i_{D1}=i_{D2}=i_{D3}=0$

Typical Design

Given: V_{CB}, I_A, $(\frac{dv_S}{dt})_{t_2\text{-}t_3}$, D_{min}, D_{max}

1. $C \geq \frac{I_A}{2(\frac{dv_S}{dt})_{t_2\text{-}t_3}}$; 2. $t_{2\text{-}3}=\frac{2CV_{CB}}{I_A}$

3. $f_s \leq \frac{1-D_{max}}{t_{2\text{-}3}}$ 4. $L=\frac{2(\frac{D_{min}}{f_s})^2}{\pi^2 C}$

5. $I_{pk}=\frac{V_{CB}}{\sqrt{\frac{2L}{C}}}$ => $\frac{I_{pk}}{I_A}=\frac{\pi \; V_{CB}}{t_{on\,min} \; (\frac{dv_S}{dt})_{t_2\text{-}t_3}}$

The diodes D_1 & D_2 must be very fast and have a very low storage charge.

APEC

Example:

Assuming: $f_s \approx 100KHz$; $t_{on\ min} \approx 1\mu S$; $\left(\frac{dv_S}{dt}\right)_{t2\text{-}t3} \approx 400V/\mu S$

$\frac{I_{pk}}{I_A} = 0.78$

✔ Check points

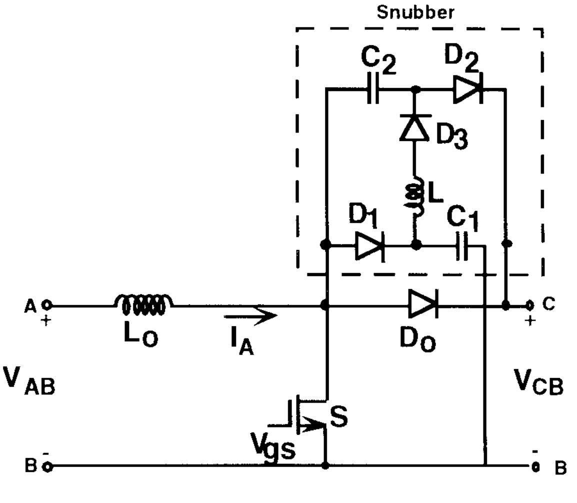
$$I_{pk}(switch) = I_A + \frac{I_A \ \pi \ V_{CB}}{t_{on\ min} \ \left(\frac{dv_S}{dt}\right)_{t2\text{-}t3}}$$

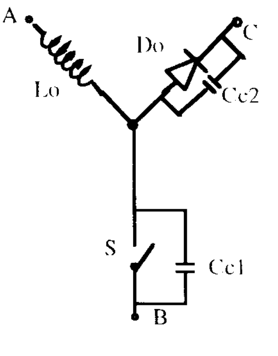
$V_{max}(switch)$ = Same as original

$V_{max}(diode)$ = Same as original

SWITCH 'TURN OFF' LOSSLESS SNUBBER

The 'One Way' Capacitor : Version 2 (SNB2) [35]

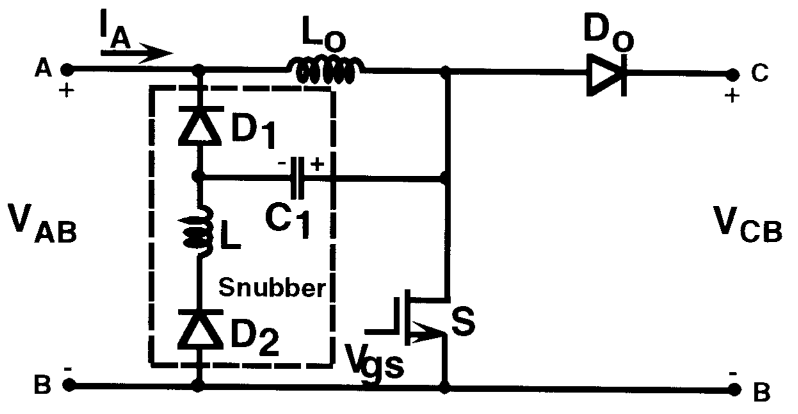


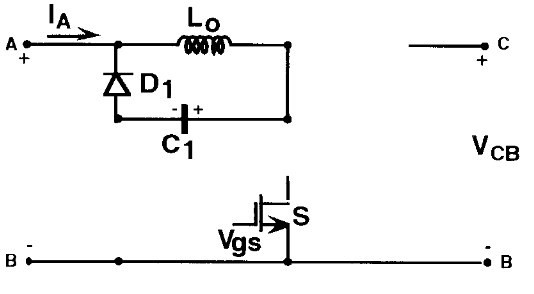


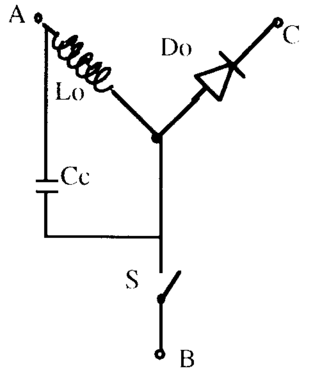
- Identical to Version 1 (normally drawn differently)

SWITCH 'TURN OFF' LOSSLESS SNUBBER (SNB3) [35]

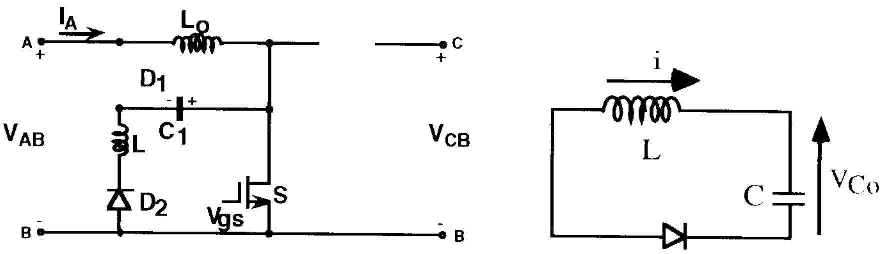
Reset to Input





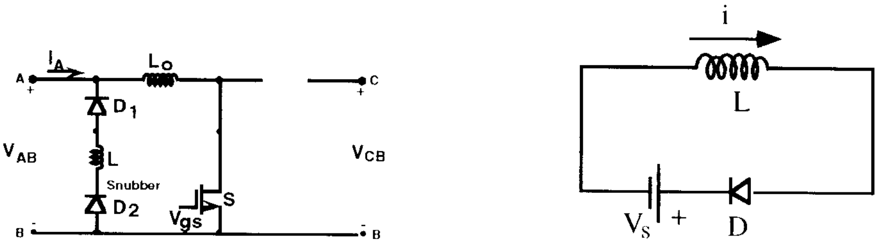


Snubbing

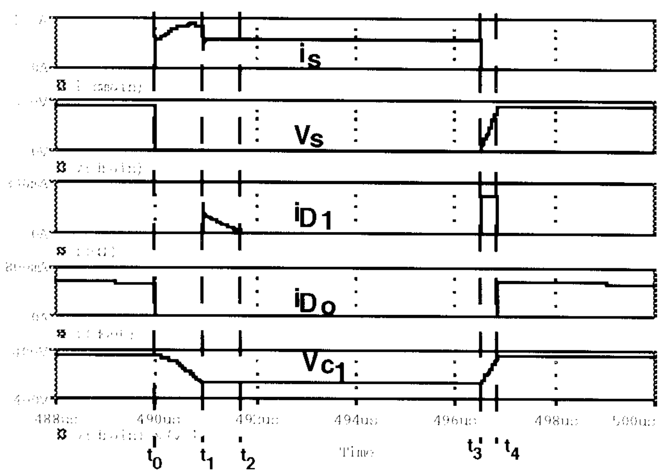


Reset 1

$V_{CA} > V_{AB}$ ---> $D_{min} = 0.5$



Reset 2



Waveforms of SNB3 (simulation)

Case $V_{CB}-V_{AB} > V_{AB}$, i.e. $V_{CB} > 2V_{AB}$

Interval t_o-t_1 $\quad i_L=i_{D2}=I_{pk}\sin(2\pi\frac{t}{T_r})$

where

$$T_r=2\pi\sqrt{LC_1}$$

$$I_{pk}=\frac{V_{CB}-V_{AB}}{\sqrt{\frac{L}{C_1}}}$$

$$i_S=I_A+i_L$$

$$v_{C1}=(V_{CB}-V_{AB})\cos(2\pi\frac{t}{T_r})$$

t_1 is found from the condition:

$$(v_{C1})_{t1}=-V_{AB}$$

$$t_1=\frac{T_r}{2\pi}\cos^{-1}(\frac{V_{AB}}{V_{CB}-V_{AB}})$$

Interval t_1-t_2 $\quad i_L=i_{D1}=i_{D2}=I_{pk}\sin(2\pi\frac{t_1}{T_r})-\frac{V_{AB}}{L}(t-t_1)$

t_2-t_1 is found from the condition: $(i_L)_{t2}=0$

$$t_2-t_1=\frac{I_{pk}L\sin(2\pi\frac{t_1}{T_r})}{V_{AB}}$$

$$i_A=I_{Ao}-i_L$$

Interval t_2-t_3 $\quad i_S=I_A \;;\; i_{Do}=i_{D1}=i_{D2}=0$

Interval t_3-t_4 $\quad i_S=i_{Do}=i_{D2}=0$

$$v_{C1}=-V_{AB}+\frac{I_{Ao}}{C_1}(t-t_3)$$

t_4-t_3 is found from the condition:

$$(v_{C1})_{t4}=V_{CB}-V_{AB}$$

$$t_4-t_3=\frac{C_1V_{CB}}{I_{Ao}}$$

Interval t_4-(t_o+T_S) $\quad i_S=i_{D1}=i_{D2}=0 \;;\; i_{Do}=I_{Ao}$

Typical Design

Given: V_{CB}, I_{Lo}, $(\frac{dv_S}{dt})_{t_3\text{-}t_4}$, $D_{min} \geq 0.5$, D_{max}

1. $C_1 \leq \frac{I_{Lo}}{(\frac{dv_S}{dt})_{t_3\text{-}t_4}}$; 2. $t_{3\text{-}4} = \frac{C_1 V_{CB}}{I_{Lo}}$; 3. $f_s \leq \frac{1-D_{max}}{t_{3\text{-}4}}$;

4. $L \approx \frac{(\frac{D_{min}}{f_s})^2}{\pi^2 C_1}$ (assumption: i_L has a sine waveform not only during interval t_o-t_1 but also during t_1-t_2)

5. $I_{pk} = \frac{V_{CB} D_{max}}{\sqrt{\frac{L}{C_1}}}$

✔ Check points

$I_{pk}(switch) = I_{Lo} + \frac{\pi D_{max} V_{CB} I_{Lo}}{t_{on\ min}(\frac{dv_S}{dt})_{t_3\text{-}t_4}}$ (no free lunch)

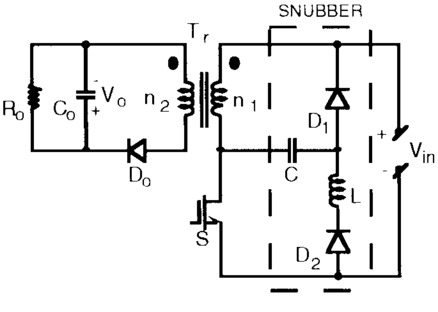
$V_{pk}(switch)$ = Same as original

$V_{pk}(diode)$ = Same as original

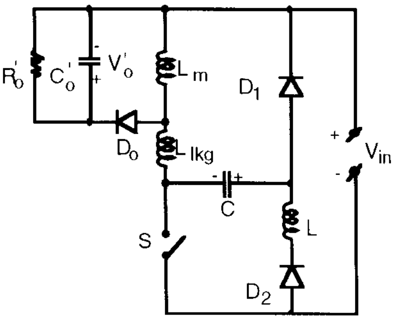
Limitation: $V_{AB} \leq \frac{V_{CB}}{2}$

In Boost Power Factor, hard switching when $V_{in} \geq \frac{V_o}{2}$

Applying snubber SNB3 in a flyback converter [6, 26, 28]



Basic topology



Equivalent circuit

L_m - output transformer magnetizing inductance

L_{lkg} - output transformer leakage inductance

C - snubber capacitance

L - snubber inductance

R_o' - reflected load resistance

C_o' - reflected capacitance of the output filter

S - switching transistor

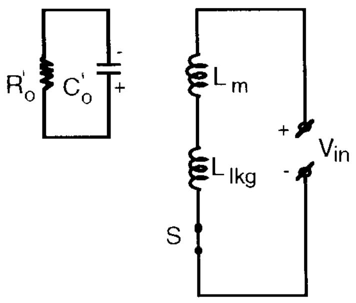
D_o - output diode

D_1, D_2- snubber diodes

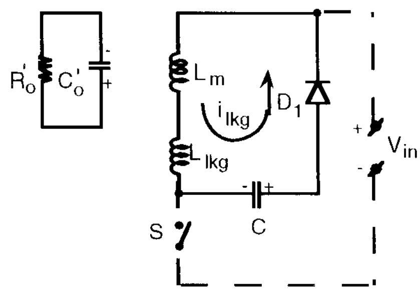
V_{in} - input voltage

V_o' - reflected output voltage

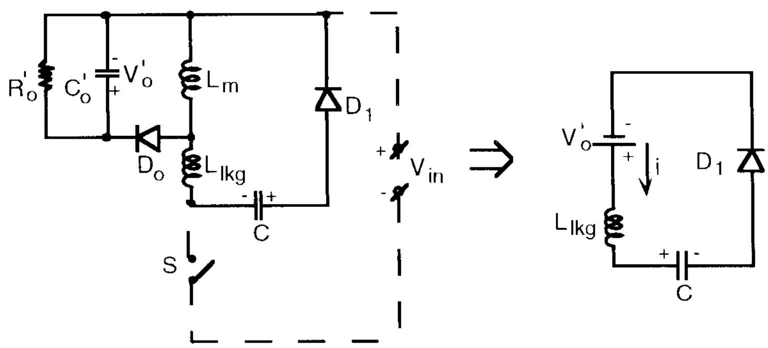
Equivalent circuit at different time intervals



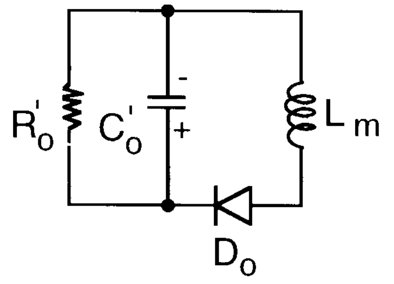
1. ON



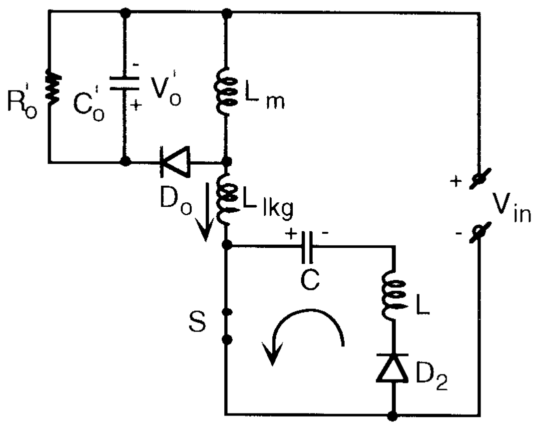
2. Snubbing 1



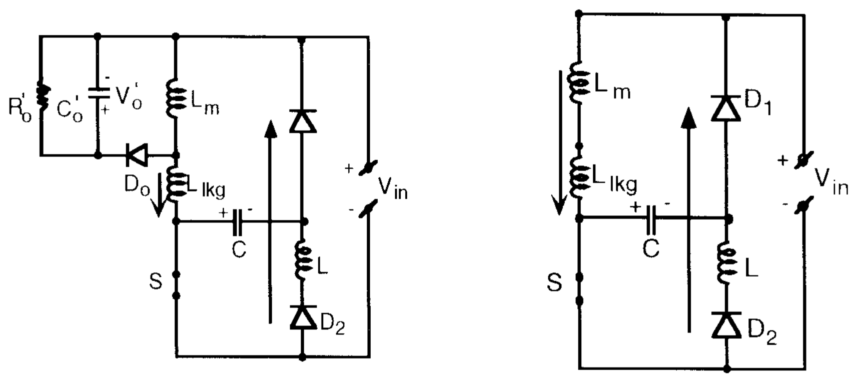
3. Snubbing 2



4. OFF



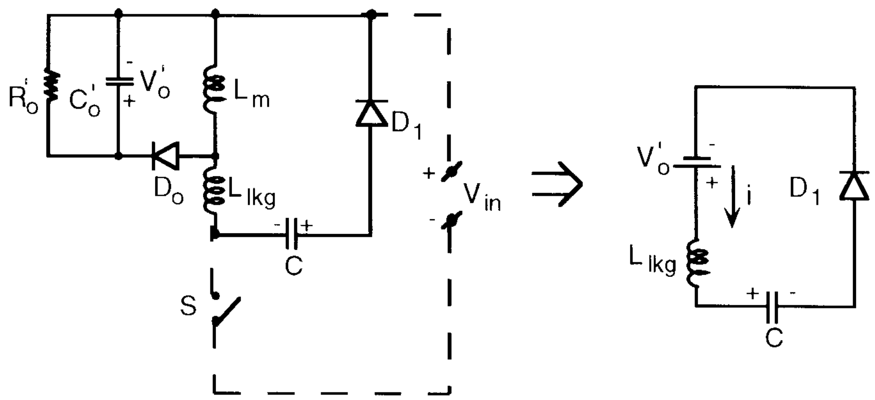
5. Resonant reset - ZCS at 'turn on' by L_{lkg} !!



6 Inductor discharge

General Observation

1. Switch maximum voltage ($V_{s\ pk}$)



Capacitor's initial voltage --> $V_C(0) = V_{in}$

Capacitor maximum voltage --> $V_{Cpk} > 2\ V_o' + V_{in}$

{ additional voltage contributed by energy removed from L_{lkg} }
Switch maximum voltage --> $V_{s\ pk} > 2\ (V_{in} + V_o')$

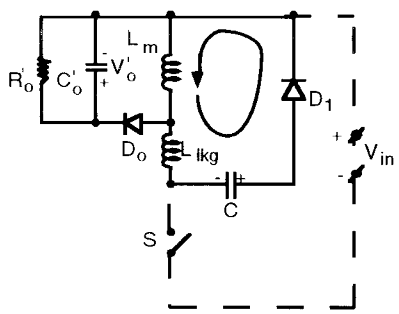
Advantages

- The snubber is simple and inexpensive [28]
- The operational duty cycle range is wider than in the case when this snubber is used in a boost converter

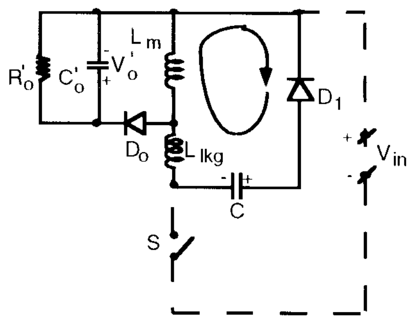
Disadvantage

- High voltage stress on the transistor $V_{s\ pk}$

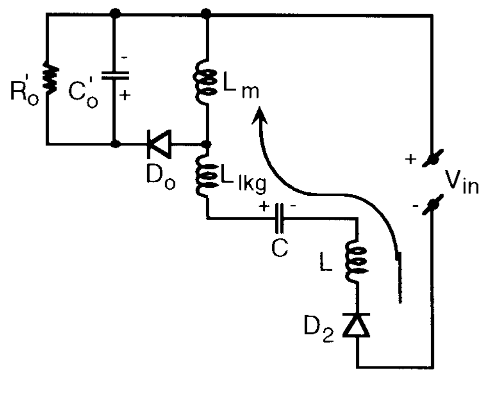
Reverse recovery of snubber diode



Resonant reset



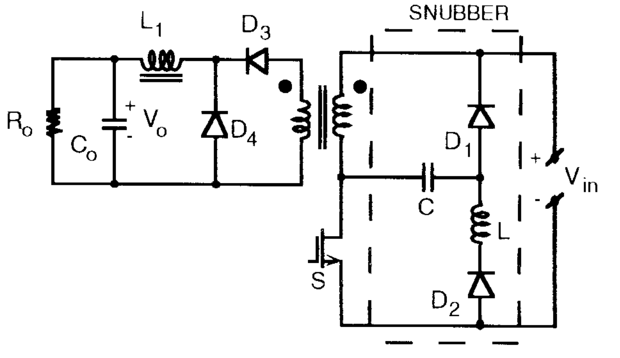
Reverse recovery



Linear discharge

- Loss of C charge
- C must retain a voltage of at least V_{in} for proper ZVS at 'turn-off'

Applying snubber SNB3 in a forward converter [40]



Same advantages and disadvantages as for the flyback converter case:

- the snubber is simple and inexpensive

- high voltage stress $V_{s\,pk}$ on transistor.

$V_{s\,pk} = V_{in}+V_{Cp}$
where

$$V_{Cp}=\sqrt{\frac{L_m I_m^2+L_{lkg} I_{Sp}^2}{C}}$$

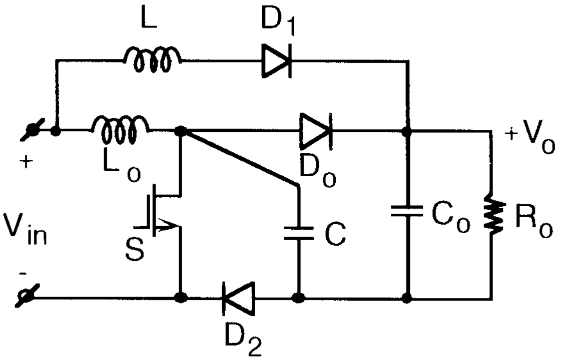
L_m - magnetizing inductance of the transformer
L_{lkg} - primary leakage inductance
I_m - magnetizing current of the transformer
$I_{s\,pk}$ - switch current in the moment when the switch is turned off

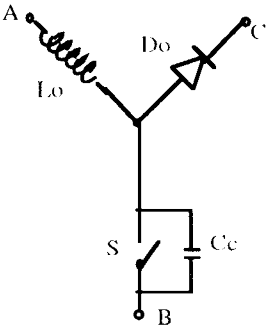
Experimental results [40]

1) V_{in}=17V; P_{out}=17W; D=2/3; $V_{s\,pk}$=52V
2) V_{in}=34V; P_{out}=36W; D=1/3; $V_{s\,pk}$=88V

APEC

A switch "turn-off" lossless snubber for a boost converter [30] (SNB4)





Snubber elements:

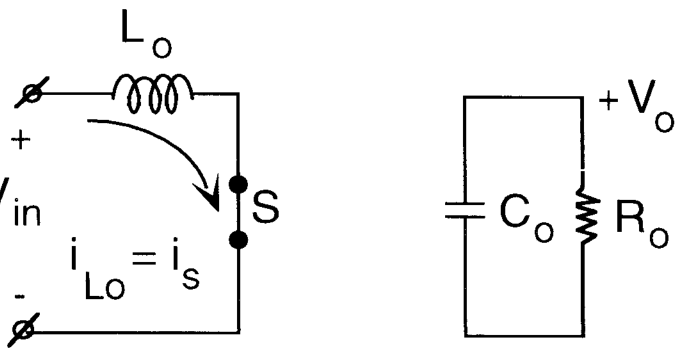
L - the resonant inductor

C - the resonant capacitor

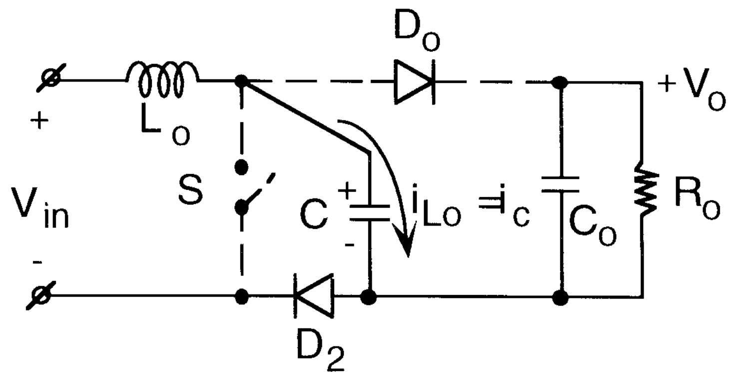
D_1, D_2 - snubber diodes

No common ground !

Equivalent circuits for different time intervals



t_1-t_2 ON

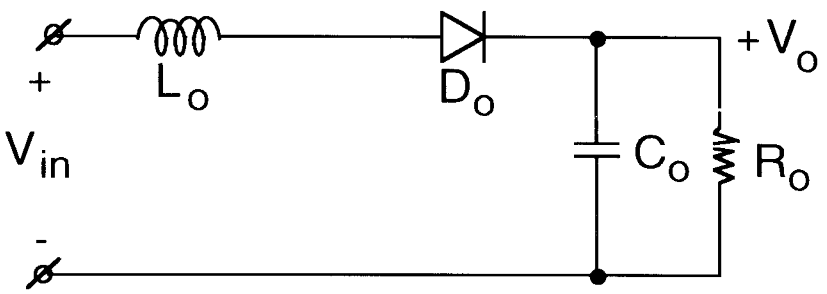


t_2-t_3 , snubbing

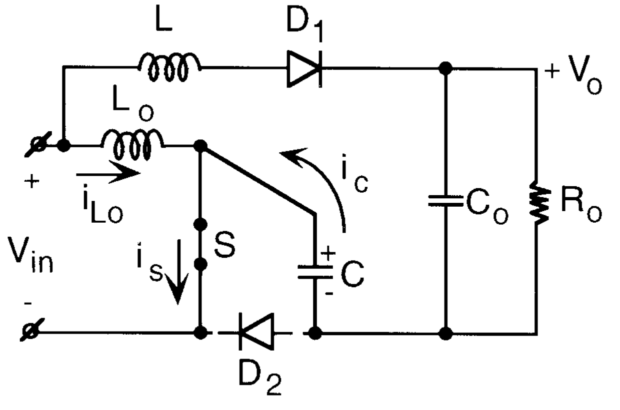
$$\frac{dv_S}{dt} = \frac{i_{Lo}}{C} \qquad \text{ZVS!}$$

$v_{Do}(t) = v_C(t)\text{-}V_o$

$v_C(t_3) = V_o \qquad v_{Do}(t_3)=0$



t_3-t_4 OFF



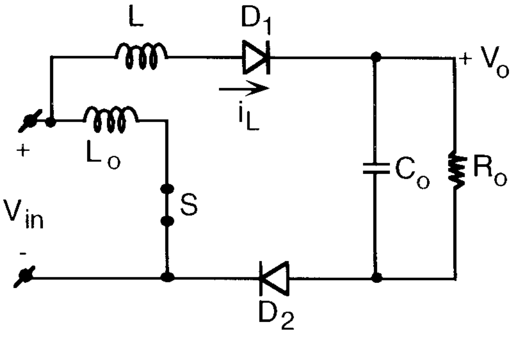
t_4-t_5 , reset

$v_C(t_4)=V_o$ $\quad$ $i_L= i_C =\frac{V_{in}}{Z_r}\sin(\omega_r t)$ $\quad$ $\omega_r= \frac{1}{\sqrt{LC}}$

$Z_r = \sqrt{\frac{L}{C}}$

$i_S=i_{Lo}+i_C$ $\quad$ $I_{Sp}=I_{Lo}+\frac{V_{in}}{Z_r}$

$v_C(t_5)=v_{D2}(t_5)=0$



t_5-t_6, linear discharge

$$i_L(t) = i_{D1}(t) = i_L(t_5) - \frac{V_o-V_{in}}{L}(t-t_5)$$

$i_L(t_6)=i_{D1}(t_6)=0$

A switch "turn-off" lossless snubber for a boost converter [30]
(SNB4)

Advantages

- Soft switching at turn off. Efficiency was reported to increase by 5% to 97% [30] .

Disadvantage

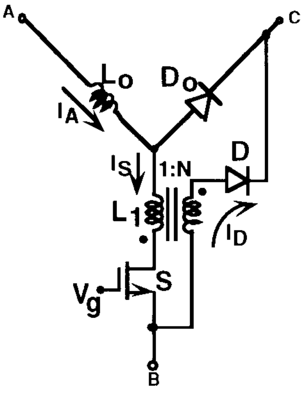
- Reverse recovery problems of diodes D_o, D_1, and D_2
- High current stress of main switch during the interval t_4 - t_5
- No common ground between input and output

Chapter 5

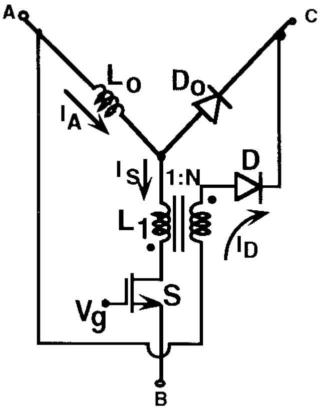
SWITCH TURN-ON AND DIODE TURN-OFF SNUBBERS

SWITCH 'TURN ON' AND DIODE 'TURN OFF' SNUBBER [35,39]

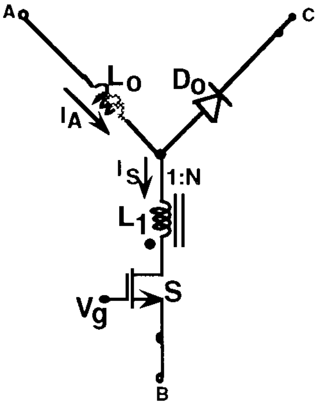
FLYBACK RESET SNUBBER (energy recovery via a catch winding)

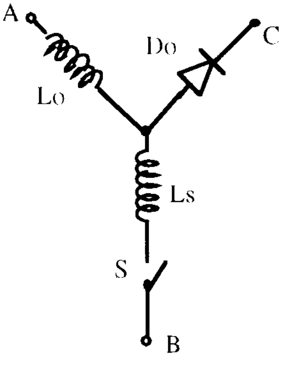


Version 1 (SNB5)

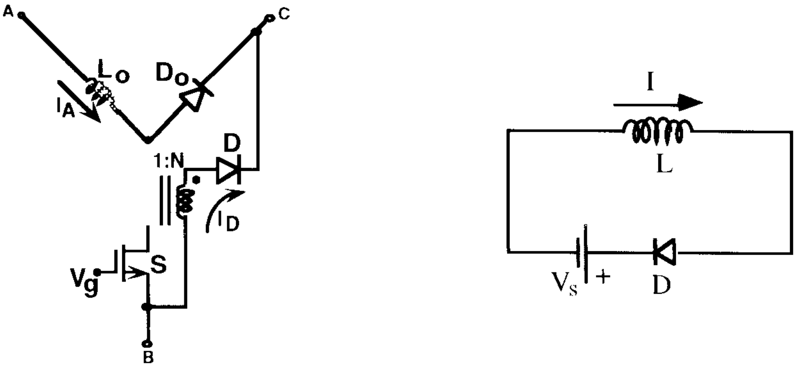


Version 2 (SNB6)



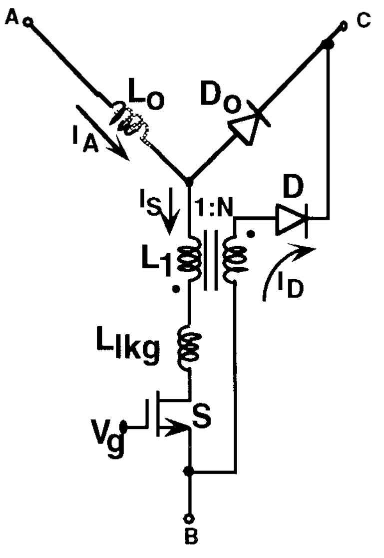


Snubbing

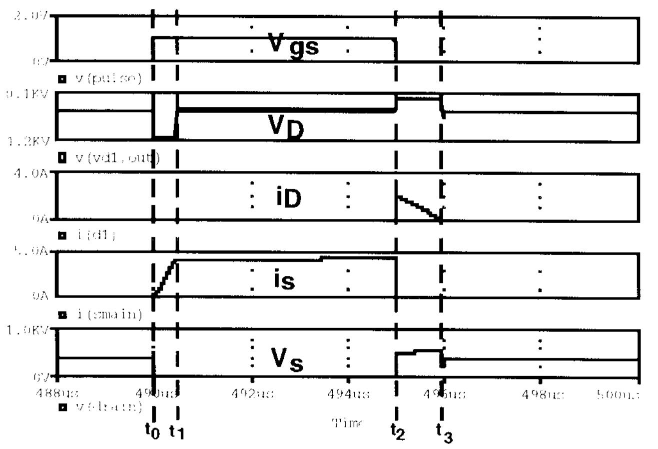


Reset

Practical Aspects (5)



- Beware of leakage inductance
- Not practical for HF high power levels



Waveforms of SNB5 (simulation)

Analysis: Version 1 (SNB5)

Interval t_o-t_1 $\qquad i_S=\frac{V_{CB}t}{L_1}$

t_1-t_o is found from the condition: $i_S(t_1)=I_A$

from where

$$t_1-t_o=\frac{L_1 I_A}{V_{CB}}$$

$$V_{Do\ max}=V_{CB}(1+N)$$

Interval t_2-t_3 $\qquad i_D=\frac{I_A}{N}-\frac{V_{CB}}{N^2 L_1}(t-t_2)$

t_3-t_2 is found from the condition: $i_D(t_3)=0$

$$t_3-t_2=\frac{I_A L_1 N}{V_{CB}} \quad ; \quad V_{S\ max}=V_{CB}\left(1+\frac{1}{N}\right)$$

$$t_3-t_2=\frac{I_A L_1 N}{V_{CB}} \quad ; \quad f_{s\ max}=\frac{1-D_{max}}{t_3-t_2}$$

$$V_{S\ max}=V_{CB}\left(1+\frac{1}{N}\right) \quad ; \quad V_{Do\ max}=V_{CB}(1+N)$$

The lower N, the shorter is t_3-t_2 and hence the higher is the upper limit of the switching frequency $f_{s\,max}$. The lower N, the higher is $V_{S\,max}$. The voltage across the main diode D_o is high when N is high.
A major disadvantage of the converter is the leakage inductance between the primary and secondary of the coupled inductor : it will cause a large voltage spike across the switch. Beware of reverse recovery problems of the auxiliary diode D.

FLYBACK RESET SNUBBER Version 1 (SNB5)

Typical Design

Given: V_{CB}, I_A, $\left(\frac{di_S}{dt}\right)_{t_o\text{-}t_1}$, D_{min}, D_{max}

1. $L_1 \leq \frac{V_{CB}}{\left(\frac{di_S}{dt}\right)_{t_o\text{-}t_1}}$; 2. $t_{1\text{-}0}=\frac{L_1 I_A}{V_{CB}}$; 3. $f_s \leq \frac{D_{min}}{t_{1\text{-}0}}$

4. $t_{2\text{-}3} \leq \frac{1-D_{max}}{f_s}$; 5. $N=\frac{V_{CB}t_{2\text{-}3}}{L_1 I_A}$; 6. $V_{S\,max}=V_{CB}(1+\frac{1}{N})$

7. $V_{Do\,max}=V_{CB}(1+N)$

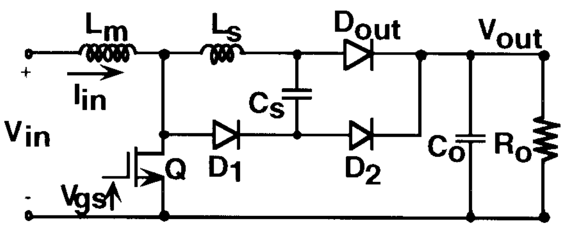
✔ Check points

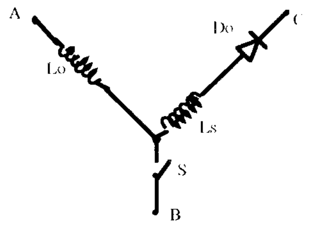
$I_{pk}(\text{switch}) = I_A$

$V_{max}(\text{switch}) = V_{CB}\,(1+\frac{1}{N}) = V_{CB}\left(1+\frac{I_A}{t_{off\,min}\left(\frac{di_s}{dt}\right)_{t0\text{-}t1}}\right)$

$V_{max}(\text{diode}) = V_{CB}\,(1+N) = V_{CB}\left(1+\frac{t_{off\,min}\left(\frac{di_s}{dt}\right)_{t0\text{-}t1}}{I_A}\right)$

LOW STRESS 'TURN ON' SNUBBER (SNB7) [17, 24]





Q - Main switch

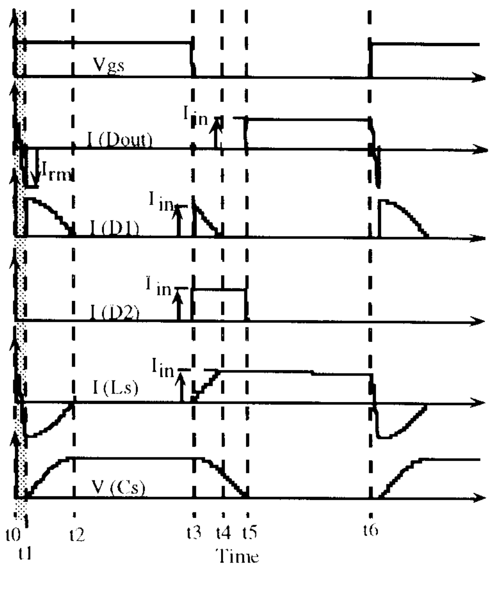
D_{out} - Main diode

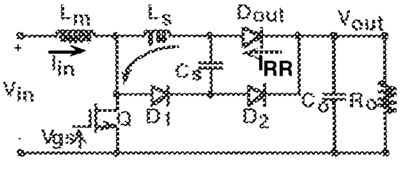
D_1,D_2 - Auxiliary diodes

L_S, C_S - Resonant network

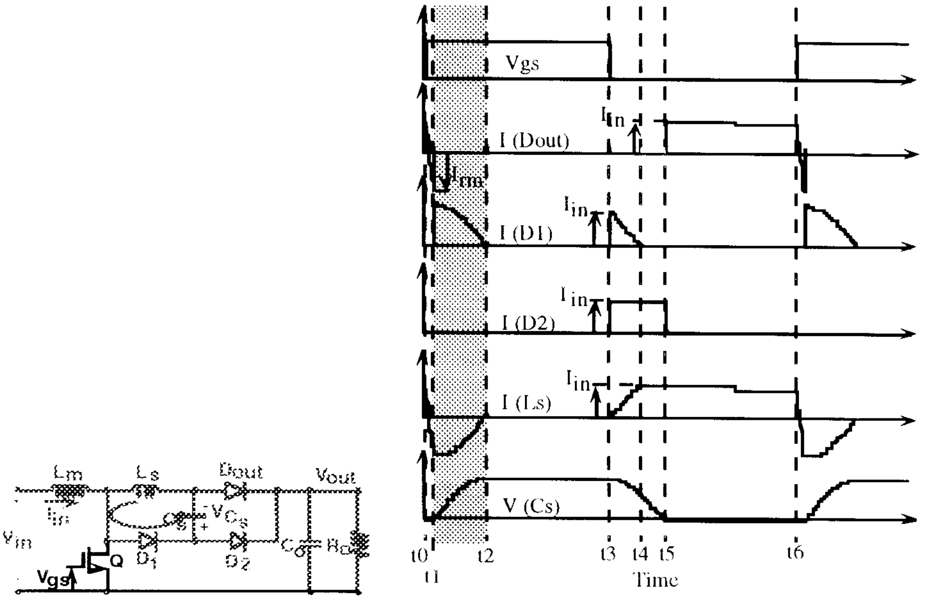
Basic waveforms of lossless snubber

Interval t_0-t_1

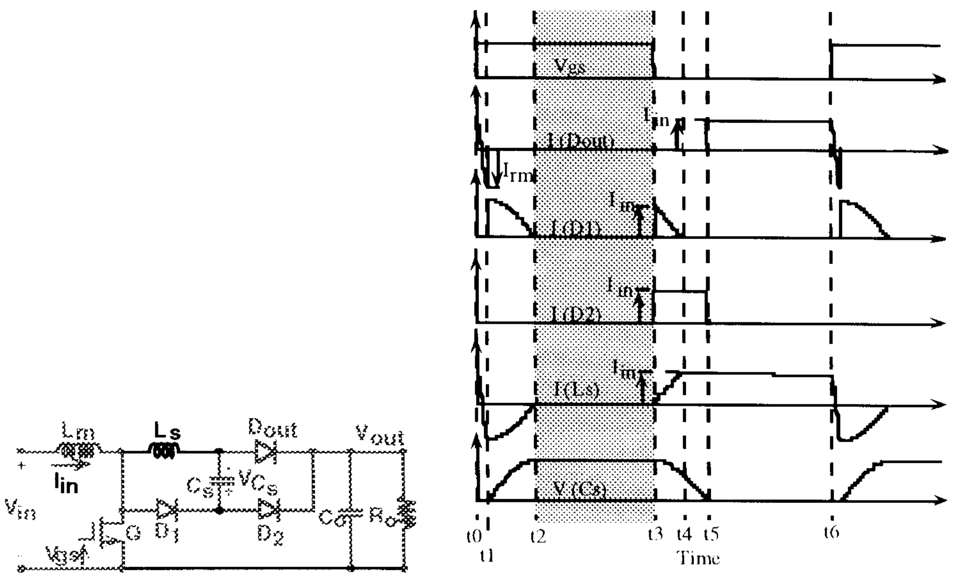




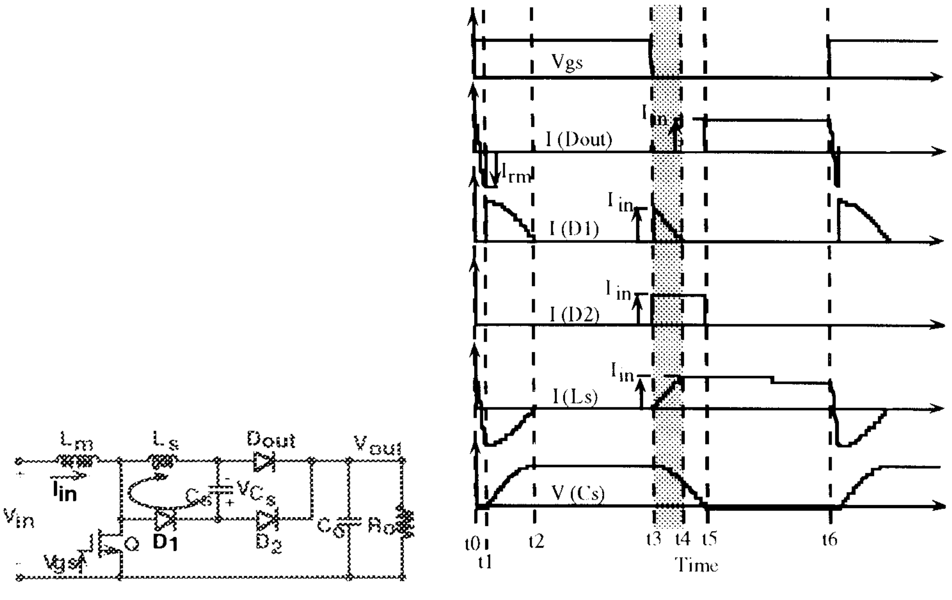
Interval t_1-t_2



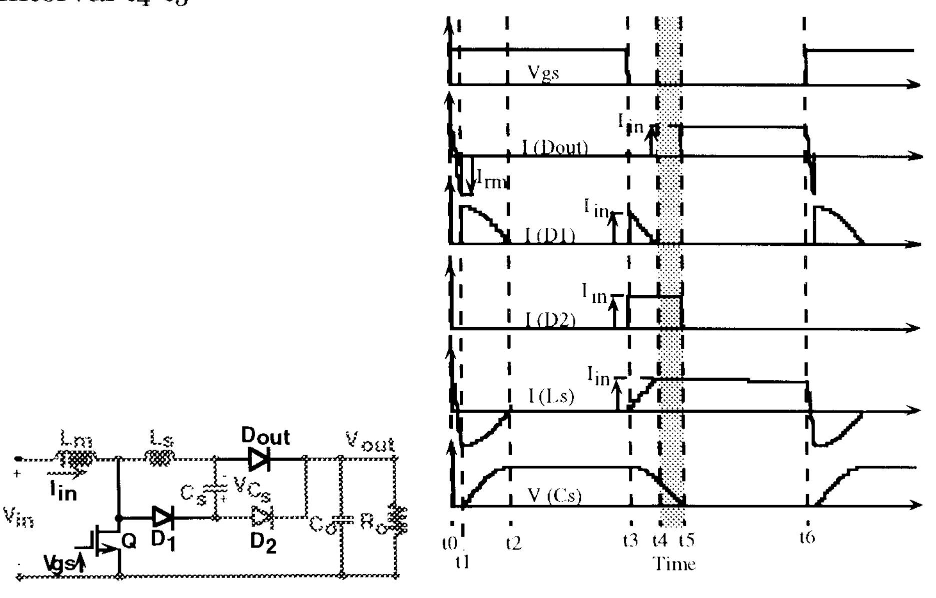
Interval t_2-t_3



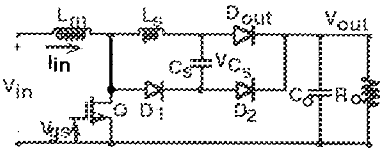
Interval t_3-t_4

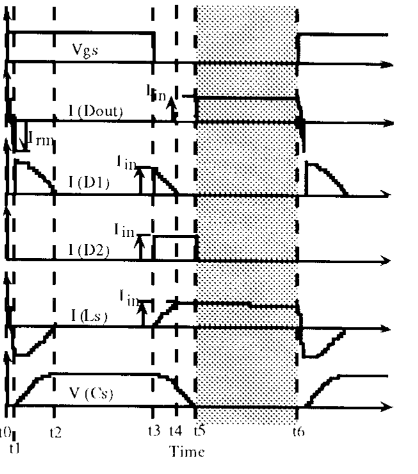


Interval t_4-t_5



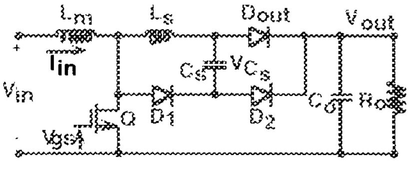
Interval t_5-t_6

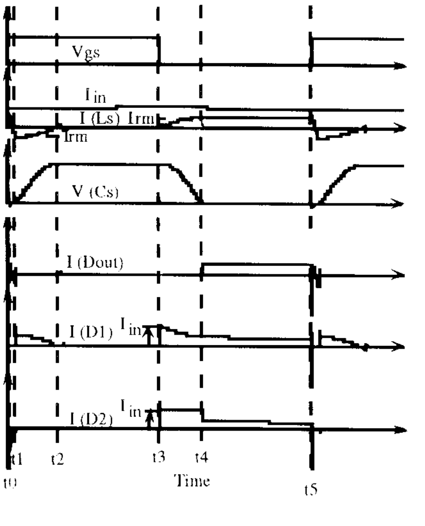




Practical requirement of peak reverse recovery current - I_{rm}

If $I_{rm} < I_{in}$

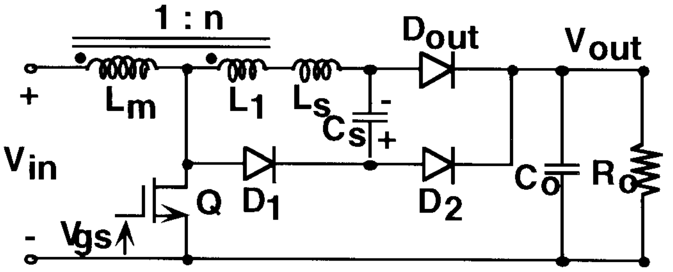




To avoid the above undesired condition

$\boxed{I_{rm} > I_{in}}$

The coupled inductor realization



D_{out} - Main diode

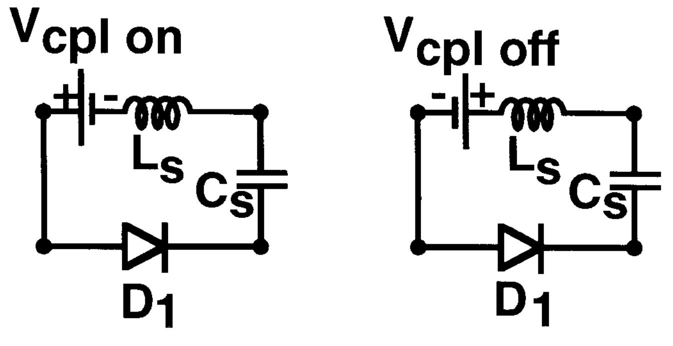
D_1, D_2 - Auxiliary diodes

L_s, C_s - Resonant network

L_m, L_1 - Coupled inductors

n - turn ratio

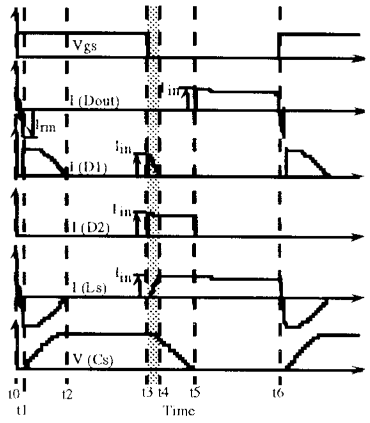
Simplified equivalent circuits of the resonant elements with coupling inductor



$V_{cpl\ on} = nV_{in}$

$V_{cpl\ off} = \frac{n}{1+n}(V_{out} - V_{in})$

Basic waveforms of the lossless snubber with coupling inductor



$V_{Cs}(t_3)$ is larger and therefore $t_{3\text{-}4}$ is smaller

Components Stress

Transistor or diode	Current stress	Current stress instance or interval	Voltage stress	Voltage stress instance or interval
Q	$I_{in}+I_{rm}$	t_1	V_{out}	t_3-t_4
D_0	I_{in} I_{rm}	t_5-t_6 t_1	$V_{out} + V_{Cmax}$	t_2
D_1	$\sqrt{I_{rm}^2 + \frac{V_{cpl\ on}^2 C_s}{L_s}}$	$t_1<t_e<t_2$	$V_{out} - V_{cpl\ on}$	t_0-t_1
D_2	I_{in}	t_3-t_5	V_{out}	t_1-t_2

$I_{rm} = \frac{(V_{out}+V_{cpl\ on})t_{rm}}{L_s}$ $\quad t_{rm} \approx t_{rr}$ $\quad t_{rr}$ = formal reverse recovery time

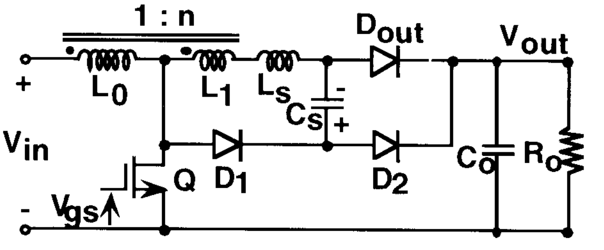
$V_{cpl\ on} = nV_{in} = n(1-D)V_{out}$

$D = \frac{t_{on}}{T_s}$

$$V_{Cmax} = I_{rm}\sqrt{\frac{L_s}{C_s}}\sin(\omega_r t_{1\text{-}2}) + V_{cpl\ on}\,[1-\cos(\omega_r t_{1\text{-}2})]$$

$$\omega_r t_{1\text{-}2} = \tan^{-1}\left(-\frac{I_{rm}\sqrt{\frac{L_s}{C_s}}}{V_{cpl\ on}}\right) + \pi$$

An experimental 1kW Boost converter



Q - IRF460

D_{out} - MUR860 t_{rr} = 60nsec.

D_1, D_2 - MUR460

L_s - 3uH C_s - 100nF

n=1/7

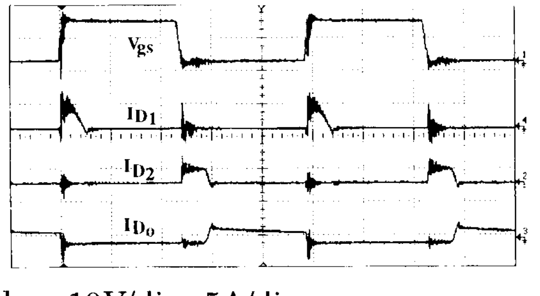
Operational conditions

F_s - 100KHz P_{out} - 1000W

V_{in} - 200V V_{out} - 400V

I_{in} - 5.8A D - 0.5

Typical waveforms of the experimental Boost converter

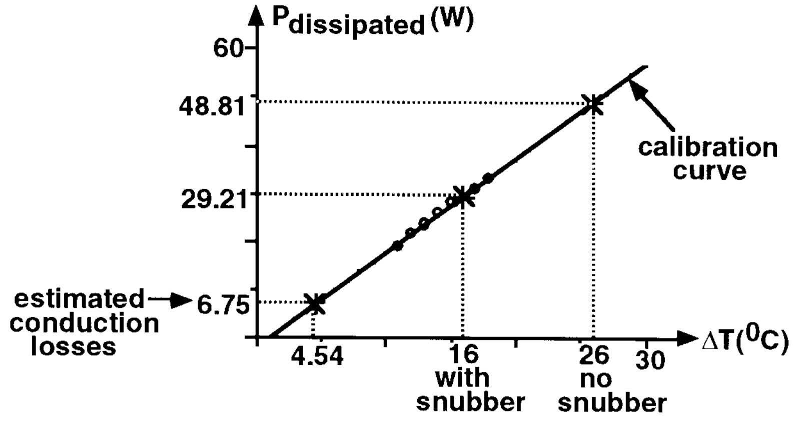


Vertical scales: 10V/div, 5A/div

Horizontal scale: 2μSec/div

✓ Smaller reverse recovery current of Do

Power dissipation



o DC calibration points

Power dissipation dropped by 19.6W

Additional Comments

✔ The main advantages of the lossless snubber:

- ZCS in turn-on
- Smaller reverse recovery current

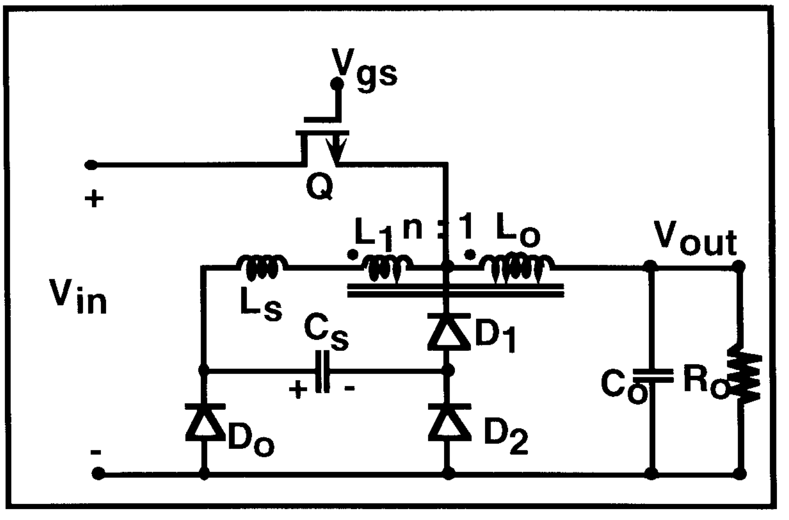
✔ The preferred embodiment is coupled inductors

✗ The main disadvantage of the snubber is limitation of the duty cycle range

✔ The overall efficiency of the experimental Boost converter was found to increase by 1%-2%

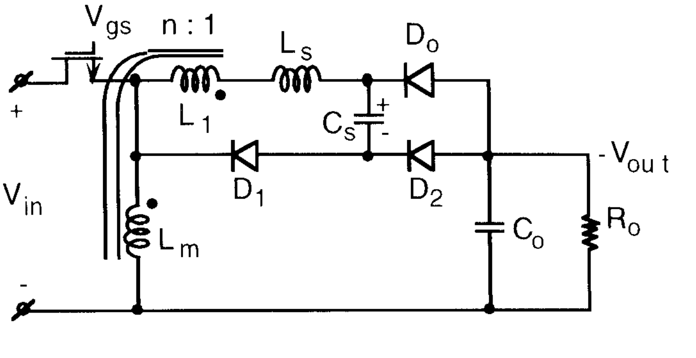
The snubber can be adopted to other PWM topologies

Implementation in Buck converter

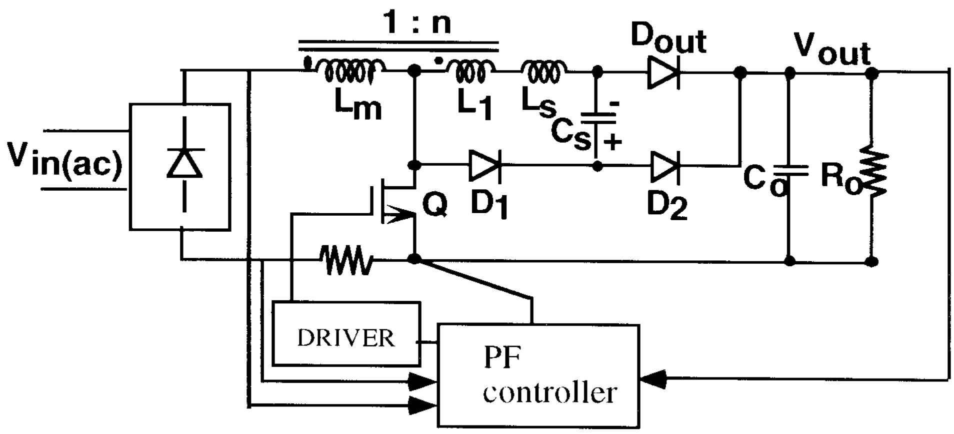


☞ Proven to increase efficiency by ≈ 2%

Implementation in a buck-boost converter



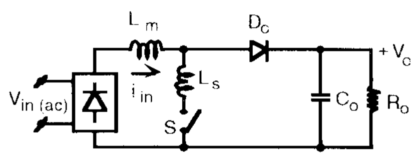
Implementation in APFC

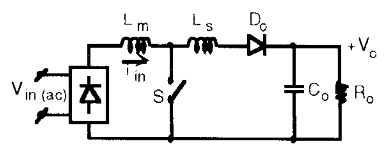


- Simple implementation
- Watch D_{onmin}
- Proven to increase efficiency by ≈ 2%

RMS current of the snubber inductor
in a boost converter implemented in APFC

Two ways of snubber inductor (L_S) connection:





(a) (b)

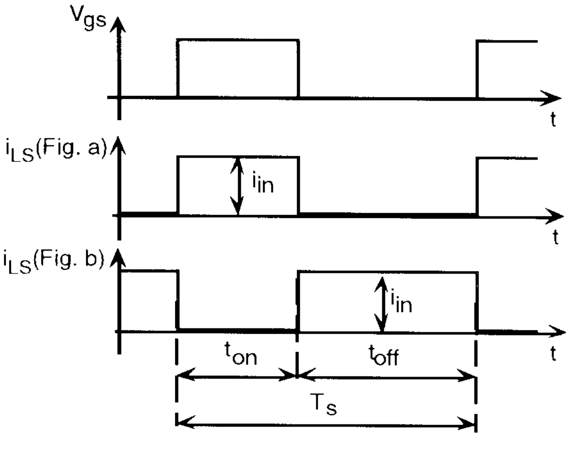
(snubber elements except L_S are not shown)

RMS current of the snubber inductor
in a boost converter implemented in APFC

Main assumptions

1. The duration of transient processes after the switch S is turned on and after S is turned off are neglected.
2. The input inductance L_m is sufficiently large that their current i_{in} is practical constant during one high frequency switching cycle T_s.

Under these conditions the current i_{Ls} of the snubber inductor has a rectangular waveform during one high frequency switching cycle T_s.



For one switching cycle T_s (high frequency period)

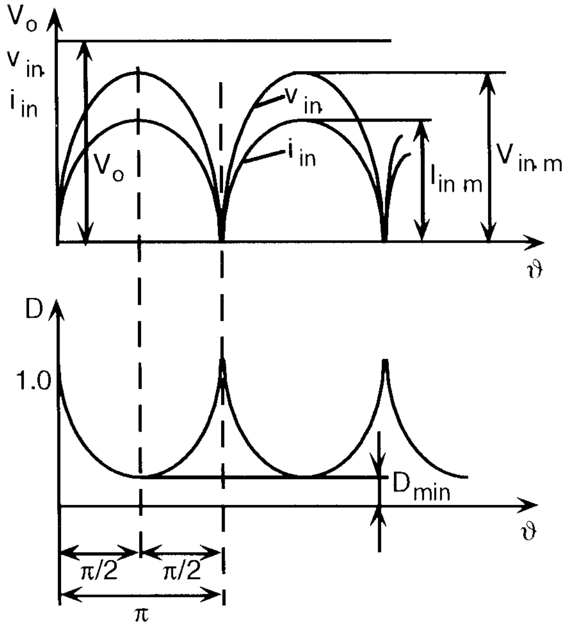
$I_{Ls\ rms\ hf}$ (Fig. a) $= i_{in}\sqrt{D}$

$I_{Ls\ rms\ hf}$ (Fig. b) $= i_{in}\sqrt{1\text{-}D}$

where $D = \dfrac{t_{on}}{T_s}$ (hf - high frequency)

RMS current of the snubber inductor
in a boost converter implemented in APFC

Ideal APFC



$v_{in} = V_{in\,m} \sin\vartheta$

$i_{in} = I_{in\,m} \sin\vartheta$

where

$\vartheta = 2\pi f_{lf} t$ (f_{lf} =50 or 60 Hz) (lf - low frequency)

V_o = const

In boost converter

$$\frac{V_o}{v_{in}} = \frac{1}{1-D} \;\; -> \;\; D=1-(1-D_{min})\sin\vartheta$$

where

$$D_{min} = 1 - \frac{V_{in\,m}}{V_o}$$

RMS current of the snubber inductor
in a boost converter implemented in APFC

Ideal APFC (cont.)

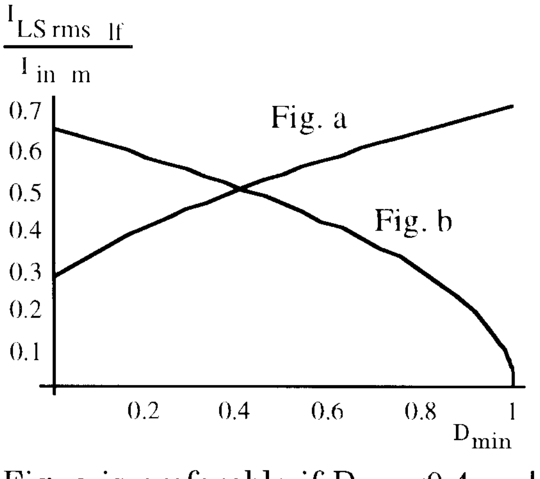
RMS current for low frequency period

$$I_{Ls\ rms\ lf\,(Fig.\ a)} = \sqrt{\frac{1}{\pi}\int_0^{\pi} I_{in\ m}^2[\sin^2\vartheta-(1-D_{min})\sin^3\vartheta]d\vartheta} =$$

$$= I_{in\ m}\sqrt{0.5-\frac{4}{3\pi}(1-D_{min})}$$

$$I_{Ls\ rms\ lf\,(Fig.\ b)} = \sqrt{\frac{1}{\pi}\int_0^{\pi} I_{in\ m}^2(1-D_{min})\sin^3\vartheta d\vartheta} =$$

$$= I_{in\ m}\sqrt{\frac{4}{3\pi}(1-D_{min})}$$



Conclusion: Fig. a is preferable if $D_{min}<0.4$ and Fig. b is preferable if $D_{min}>0.4$.

Chapter 6

TURN-ON AND TURN-OFF SINGLE SWITCH SNUBBERS

TURN-ON AND TURN-OFF SINGLE SWITCH SNUBBERS

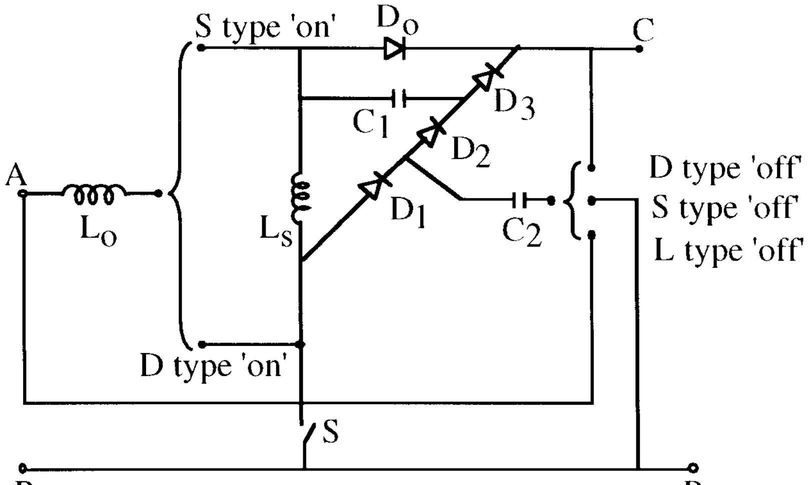
Possible Realizations:

LCC - one resonant inductor and two resonant capacitors
[10, 16, 22, 31, 32, 34, 39, 46]

LLC -two resonant inductors and one resonant capacitor
[1, 23, 29, 45]

LC - one resonant inductor and one resonant capacitor
[2, 22, 36, 41]

Generic LCC 'turn-off' & 'turn-on' snubber topology
(see also [32])

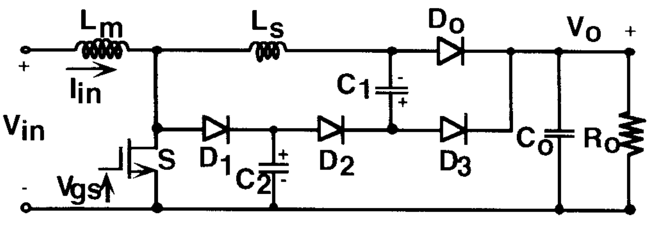


- Snubber analysis is identical for all connections

TURN-ON AND TURN-OFF SINGLE SWITCH SNUBBERS

Boost converter with LCC snubber: D type 'on' and S type 'off' (SNB8) [22, 31, 32]

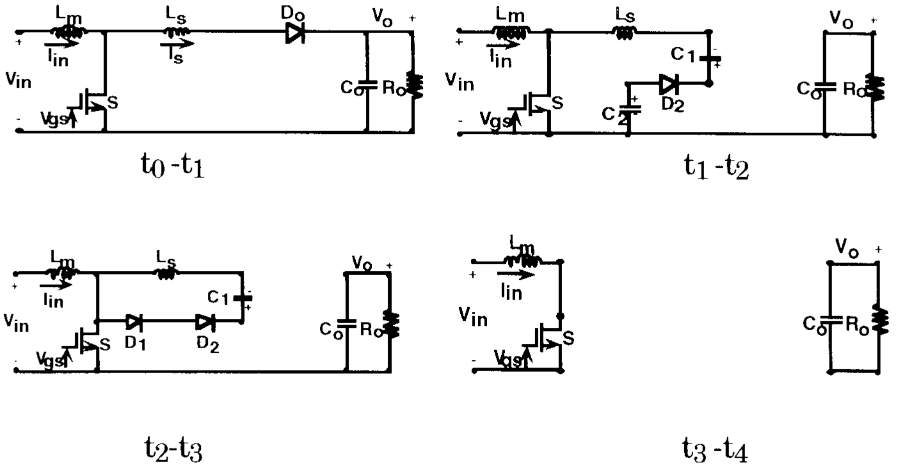
(Modified turn-on snubber SNB7)

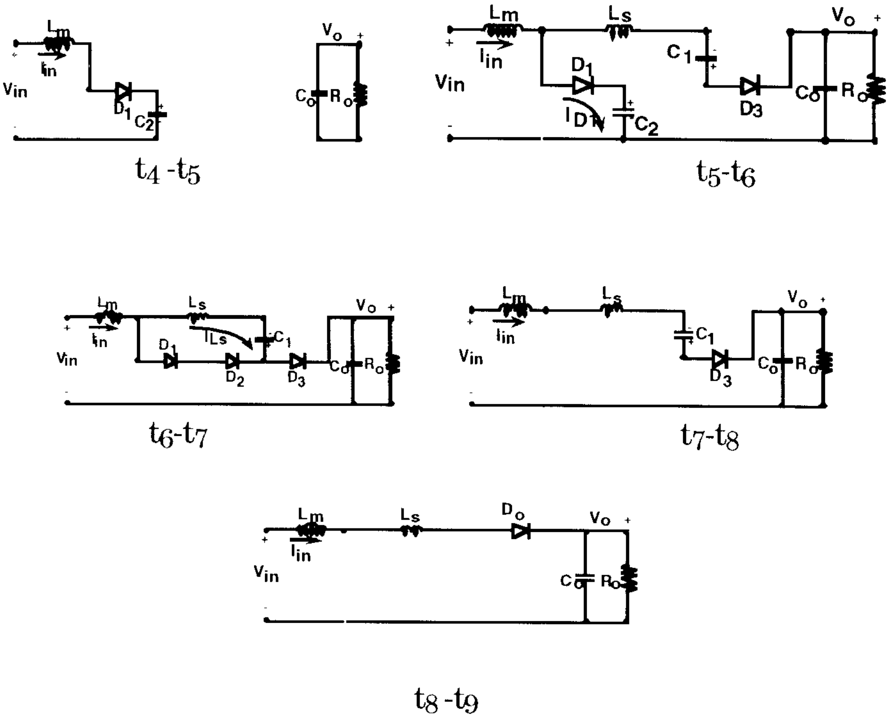


S - main switch

D_0 - main diode

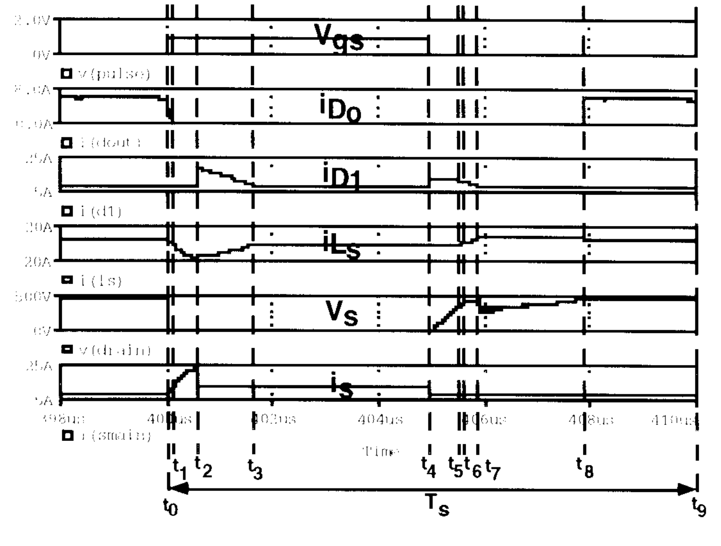
D_1, D_2, D_3 - auxiliary diodes





L_m
L_s
I_{in}
V_{in}
D_1
D_2
D_3
D_o
C_1
C_2
C_o
R_o
V_o
I_{D1}
I_{Ls}
t4 -t5
t5-t6
t6-t7
t7-t8
t8 -t9

Simulation



Vgs
iDo
iD1
iLs
Vs
is
t1
t2
t3
t4
t5
t6
t7
t8
t0
Ts
t9
Time

Interval t_0-t_1

$$i_s=\frac{V_o}{L_s}t \qquad \text{ZCS!}$$

$$\frac{di_s}{dt}=\frac{V_o}{L_s}$$

$$i_{Do}=I_{in}-i_s$$

t_1-t_o is found from the condition: $i_{Do}(t_1)=0$

$$t_1-t_o=\frac{I_{in}L_s}{V_o}$$

$$v_{C1}=0 \quad ; \quad v_{C2}=V_o$$

Interval t_1-t_2 $\qquad i_s=I_{in}+i_{Ls}$

$$i_{Ls}=i_{C1}=i_{C2}=i_{D2}=\frac{V_o}{\sqrt{\frac{L_s}{C_{12}}}}\sin[\omega_{r12}(t-t_1)]$$

where

$$C_{12}=\frac{C_1C_2}{C_1+C_2}\ ; \qquad \omega_{r12}=\frac{1}{\sqrt{L_sC_{12}}}$$

$$v_{C1}=V_o\frac{C_2}{C_1+C_2}\{1-\cos[\omega_{r12}(t-t_1)]\}$$

$$v_{C2}=-v_{D1}=V_o\frac{C_2}{C_1+C_2}\{1+\frac{C_1}{C_2}\cos[\omega_{r12}(t-t_1)]\}$$

t_2-t_1 is found from the condition: $v_{D1}(t_2)=0$

$$t_2-t_1=\frac{1}{\omega_{r12}}\cos^{-1}(-\frac{C_2}{C_1})=\sqrt{L_sC_{12}}\cos^{-1}(-\frac{C_2}{C_1})$$

$$i_{Do}=i_{D1}=i_{D3}=0$$

Interval t_2-t_3 $i_s=I_{in}$

$$i_{Ls}=i_{C1}=i_{D1}=i_{D2}=-\frac{V_{C1}(t_2)}{\sqrt{\frac{L_s}{C_1}}}\sin[\omega_{r1}(t-t_2)]+i_{Ls}(t_2)\cos[\omega_{r1}(t-t_2)]$$

where

$$\omega_{r1}=\frac{1}{\sqrt{L_sC_1}}$$

t_3-t_2 is found from the condition: $i_{D1}(t_3)=i_{D2}(t_3)=0$

$$t_3-t_2=\frac{1}{\omega_{r1}}\tan^{-1}\left(\frac{i_{Ls}(t_2)\sqrt{\frac{L_s}{C_1}}}{v_{C1}(t_2)}\right)=\frac{1}{\omega_{r1}}\tan^{-1}\left(\sqrt{\frac{C_1-C_2}{C_2}}\right)$$

$$v_{C1}=v_{C1}(t_2)\cos[\omega_{r1}(t-t_2)]+i_{Ls}(t_2)\sqrt{\frac{L_s}{C_1}}\sin[\omega_{r1}(t-t_2)]=$$

$$=V_o\frac{C_2}{C_1}\cos[\omega_{r1}(t-t_2)]+V_o\frac{\sqrt{C_2(C_1-C_2)}}{C_1}\sin[\omega_{r1}(t-t_2)]$$

$$v_{C1}(t_3)=V_o\sqrt{\frac{C_2}{C_1}}$$

Interval t_3-t_4 $i_s=I_{in}$

$$i_{Do}=i_{D1}=i_{D2}=i_{D3}=0$$

$$v_{Do}=-V_o$$

$$v_{C1}=v_{C1}(t_3)=V_o\sqrt{\frac{C_2}{C_1}} \qquad (C_2<C_1)$$

$$v_{C2}=0$$

Interval t_4-t_5 $i_s=0$

$$i_{D1}=I_{in}$$

$$v_s=v_{C2}=\frac{I_{in}(t-t_4)}{C_2} \qquad \frac{dv_s}{dt}=\frac{I_{in}}{C_2} \qquad \text{ZVS!}$$

$$C1=v_{C1}(t_3)=V_o\sqrt{\frac{C_2}{C_1}} \qquad v_{D3}=v_{C2}+v_{C1}-V_o$$

t_5-t_4 is found from the condition: $v_{D3}(t_5)=0$

$$t_5-t_4=\frac{V_oC_2\left(1-\sqrt{\frac{C_2}{C_1}}\right)}{I_{in}}$$

Interval t_5-t_6

$i_s=0; \quad i_{Do}=i_{D2}=0$

$$i_{D1}=I_{in}\frac{C_1}{C_1+C_2}\{\frac{C_2}{C_1}+\cos[\omega_{r12}(t-t_5)]\}$$

$$v_{C2}=V_o(1-\sqrt{\frac{C_2}{C_1}})+\frac{I_{in}}{\omega_{r12}(C_1+C_2)}\{\omega_{r12}(t-t_5)+\frac{C_1}{C_2}\sin[\omega_{r12}(t-t_5)]\}$$

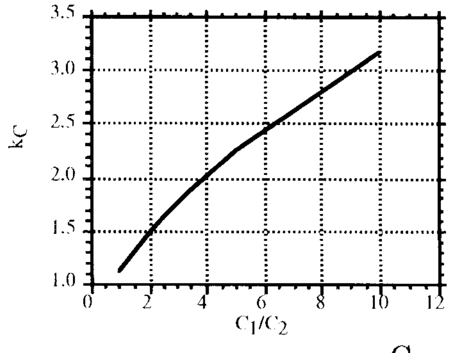
Two operating modes are possible. They depend on the condition at t_6 of the interval t_5-t_6:

Mode 1 will prevail if : $(v_{C2})_{t6} = V_o$ $\quad (i_{D1})_{t6} > 0$

Mode 2 will prevail if : $(i_{D1})_{t6} = 0$ $\quad (v_{C2})_{t6} < V_o$

Necessary condition for Mode 1: $\frac{V_o}{I_{in}}\sqrt{\frac{C_2}{L_s}} < k_C$

where : $$k_C=\frac{1}{(1+\frac{C_2}{C_1})\sqrt{\frac{C_1}{C_2}+1}}[\cos^{-1}(-\frac{C_2}{C_1})+\sqrt{(\frac{C_1}{C_2})^2-1}]$$



The dependence of k_C on $\frac{C_1}{C_2}$

- For Mode 1: $\frac{V_o}{I_{in}}\sqrt{\frac{C_2}{L_s}} < k_C$

t_6-t_5 is found from the condition: $v_{D2}(t_6)=0$; $v_{C2}(t_6)=V_o$

Interval t_6-t_7 , Mode 1, $i_s=0$; $v_s=V_o$

$i_{D3}=I_{in}$

$i_{D1}=i_{D2}=I_{in}-i_{Ls}$

$$i_{Ls}=\frac{V_{C1}(t_3)}{\sqrt{\frac{L_s}{C_1}}}\sin[\omega_{r1}(t-t_6)]$$

t_7-t_6 is found from the condition: $i_{D1}(t_7)=i_{D2}(t_7)=0$

$$t_7-t_6=\frac{1}{\omega_{r1}}\sin^{-1}\left(\frac{I_{in}\sqrt{\frac{L_s}{C_1}}}{v_{C1}(t_3)}\right)=\frac{1}{\omega_{r1}}\sin^{-1}\left(\frac{I_{in}\sqrt{\frac{L_s}{C_2}}}{V_o}\right)$$

Interval t_7-t_8 , Mode 1, $i_s=i_{Do}=i_{D1}=i_{D2}=0$; $i_{D3}=I_{in}$

$$v_{C1}=v_{C1}(t_7)-\frac{I_{in}(t-t_7)}{C_1}$$

t_8-t_7 is found from the condition: $v_{C1}(t_8)=0$

$$t_8-t_7=\frac{v_{C1}(t_7)C_1}{I_{in}}$$

$v_s \approx V_o-v_{C1}$

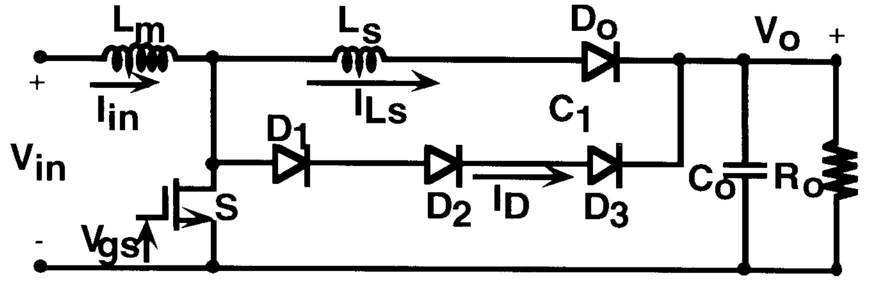
Interval t_8-t_9 , Mode 1, $i_s=i_{D1}=i_{D2}=i_{D3}=0$;

$i_{Do}=I_{in}$

$v_s \approx V_o$; $v_{C1}=0$; $v_{C2}=V_o$

LOSSLESS TURN-ON AND TURN-OFF SNUBBER (SNB8)

Mode 2. Time interval before the switch Q is turned on



- Three auxiliary diodes are conducting
- Hard switching
- Coupling will help to avoid Mode 2.

TURN-ON TURN-OFF SNUBBER (SNB8)

Typical Design

Given: V_o, I_{in}, $(\frac{di_S}{dt})_{t_0\text{-}t_1}$, $(\frac{dv_S}{dt})_{t_4\text{-}t_5}$, D_{min}, D_{max}

1. $L_S \geq \frac{V_o}{(\frac{di_S}{dt})_{t_0\text{-}t_1}}$

2. $C_2 \geq \frac{I_{in}}{(\frac{dv_S}{dt})_{t_4\text{-}t_5}}$

3. $k_C \quad > \frac{V_o}{I_{in}}\sqrt{\frac{C_2}{L_s}} \quad > 1$

4. Select C_1/C_2 using the plot $k_C=\varphi(\frac{C_1}{C_2})$.

5. $t_{on\ min}=\frac{I_{in}L_s}{V_o} + \sqrt{L_sC_{12}}\cos^{-1}(-\frac{C_2}{C_1}) +\sqrt{L_sC_1}\tan^{-1}(\sqrt{\frac{C_1-C_2}{C_2}})$

6. $t_{off\ min} \approx \frac{V_oC_2}{I_{in}}[1+\sqrt{\frac{C_1}{C_2}(1-\frac{I_{in}^2L_s}{V_o^2C_2})}] + \sqrt{L_sC_1}\sin^{-1}(\frac{I_{in}\sqrt{\frac{L_s}{C_2}}}{V_o})$

7. $D_{min} = (t_{on\ min}) f_s$; $D_{max} =(1 - t_{off\ min}) f_s$

8. Current stress of the switch S

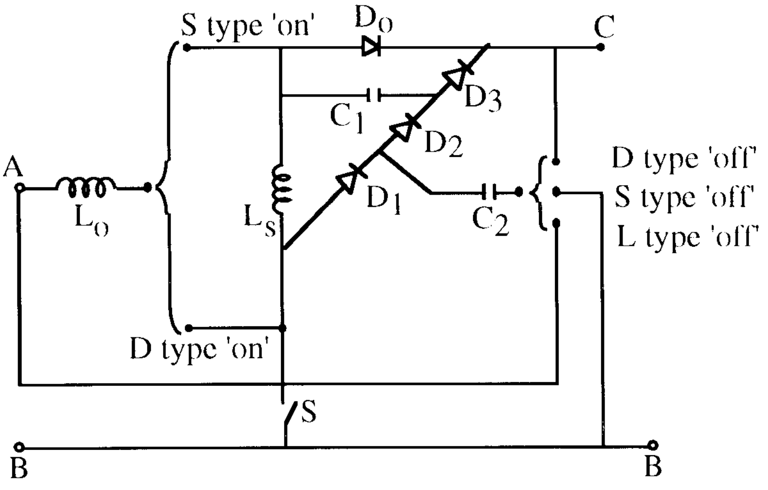
$$I_{s\ pk}=I_{in}+V_o\sqrt{\frac{C_{12}}{L_s}}$$

9. Voltage stress of the switch S

$$V_{s\ pk} = V_o$$

Apply iteration to approach desired design goals

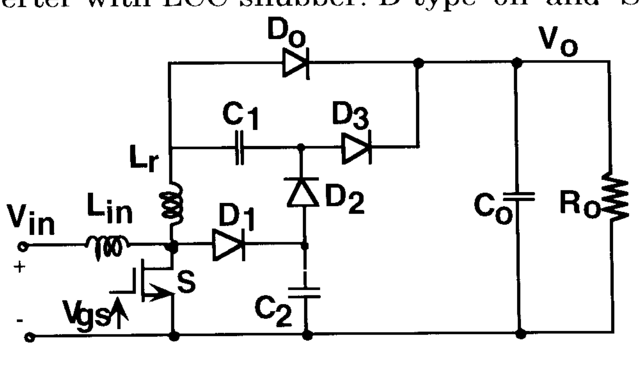
Some Additional Recent Turn-On & Turn-Off Passive Lossless LCC Snubbers



- Generic LCC 'turn-off' & 'turn-on' snubber topology

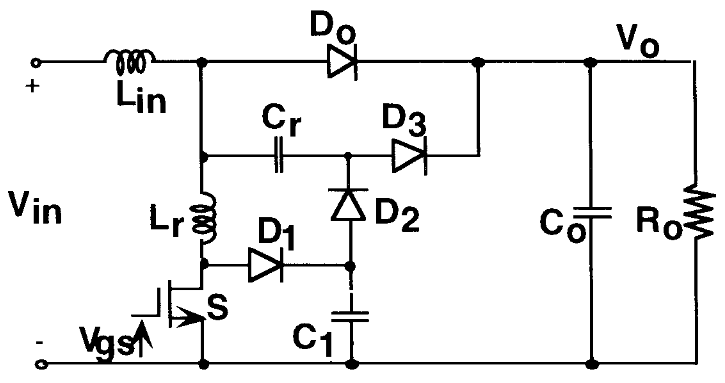
SNB8 discussed above (drawn differently)

Boost converter with LCC snubber: D type 'on' and S type 'off'

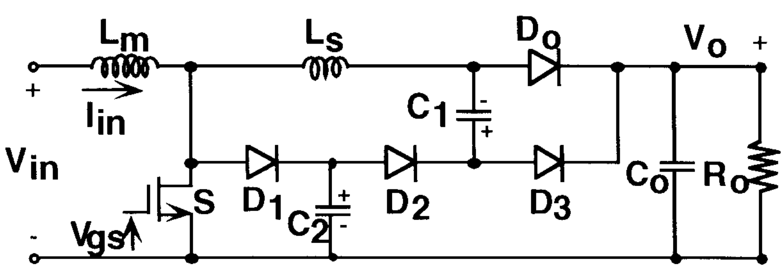


Moving the connection point to main inductor

Boost converter with LCC snubber: S type 'on'
and S type 'off' (SNB9) [16, 32, 39]

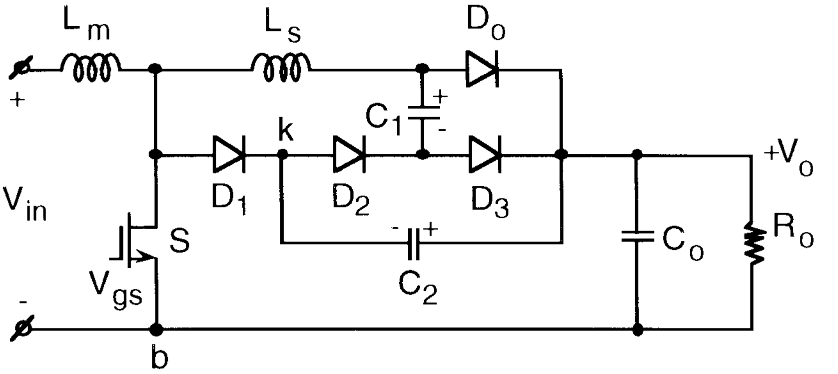


Topologies similar to SNB8 and SNB9
with different capacitor C_2 connections [31, 32, 34, 46]



Original

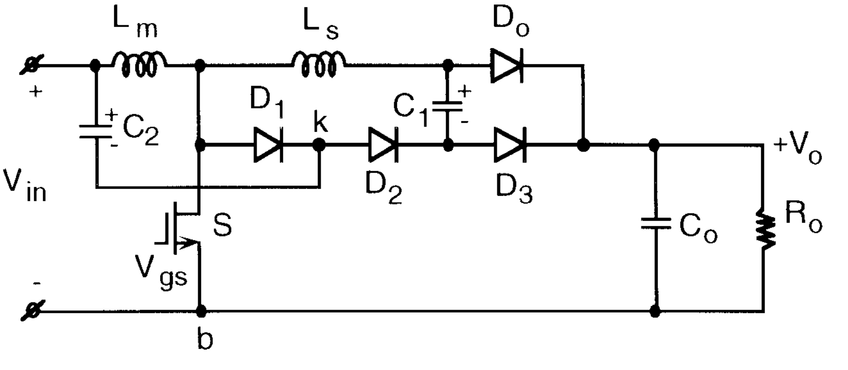
SNB10: D type 'on' and D type 'off' [34, 46]



C_2 connected to output

Ripple to output

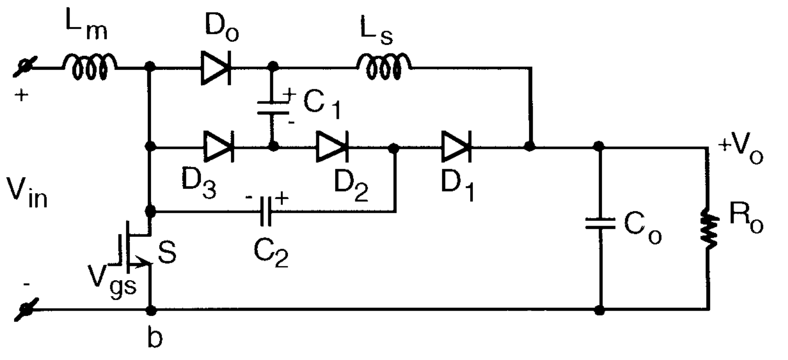
SNB11: D type 'on' and L type 'off' [32]



C_2 connected to input

Ripple stired to input

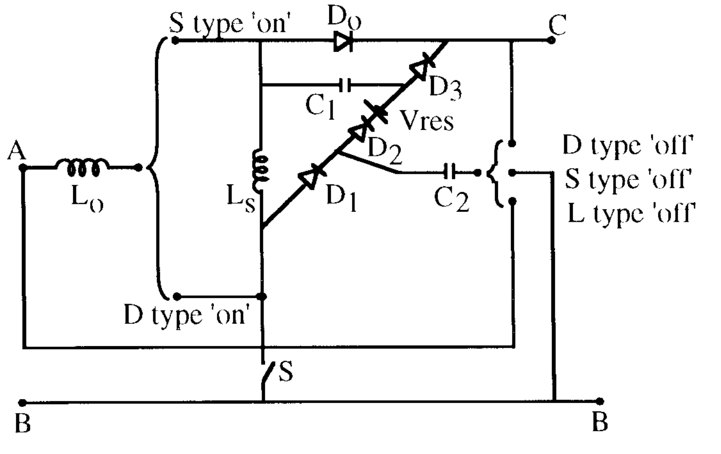
SNB12: D type 'on' and L type 'off' [31]



C_2 connected to switch

Stirs ripple to output

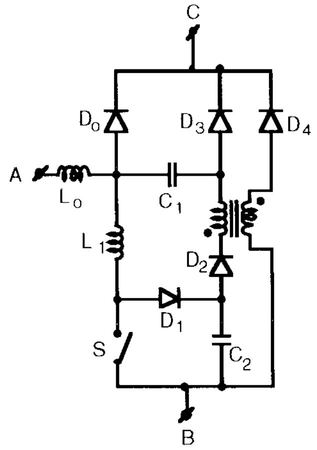
Generic LCC 'turn-off' & 'turn-on' snubber topology with improved reset



- V_{res} accelerates the reset phase

Turn-on and turn-off snubber for single switch converter

Improved Williams snubber SNB13 [10]



Operating process is based on a dual energy recovery circuit, improvement on SNB9.

- V_{res} replaced by transformer

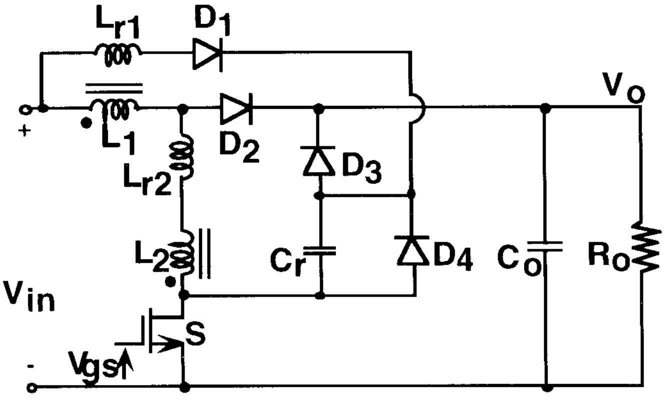
Claimed to have:

1. Lower peak current in the switch S.
2. Shorter discharge time is needed for the capacitor C_1 at low load . Hence, wider operation range of load current and higher switching frequency.

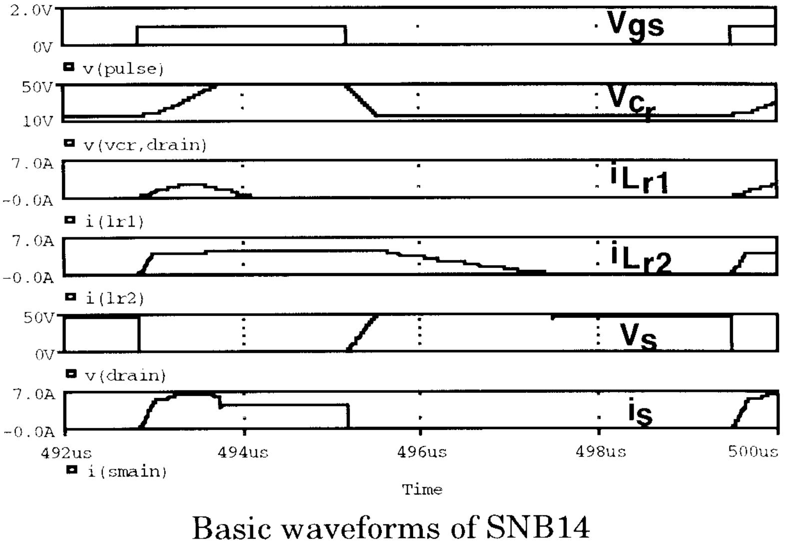
Reset through transformer (leakage and reset problems ?)
See also snubbers with recovery transformers [11-13]

Boost converter with LLC-type snubber (SNB14)

[1, 23, 45]



Coupled inductor



Basic waveforms of SNB14

SNB14 Features

Pro

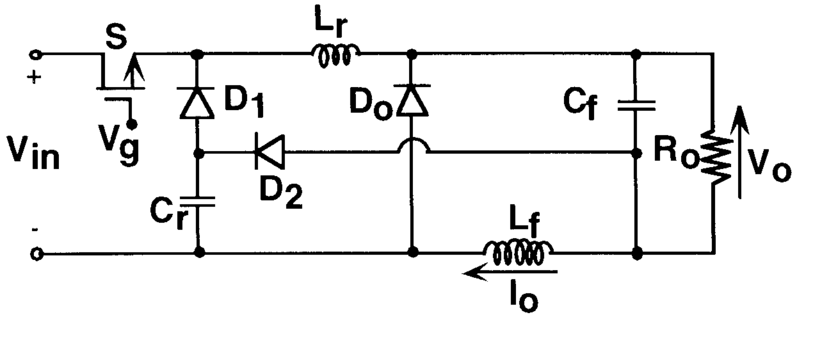
1. Soft switching at turn-on and turn-off in a wide load range without high current and voltage stresses.
2. Coupling decreases the minimal needed value of turn-off time of the switch.

Con

1. The resonant circuit L_1C_r is fed from the input voltage V_{in}. Hence, the peak capacitor's voltage cannot be higher $2V_{in}$. Consequently, when $V_o > 2V_{in}$ => hard switching.

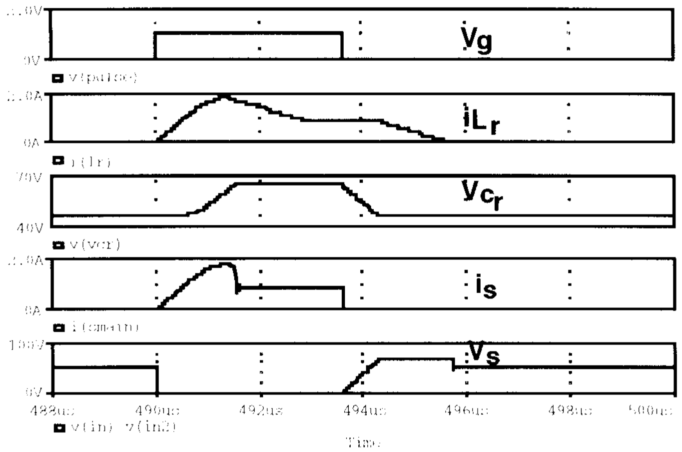
Buck converter with LC type snubber (SNB15) [2, 36]

(one inductor and one capacitor)



- No common ground !

See boost converter with LC type snubber in [41]

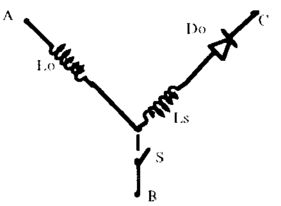


Waveforms of SNB15

{'Turn-on'}, {'turn-off'} and {'turn-on' - 'turn-off'} snubbers

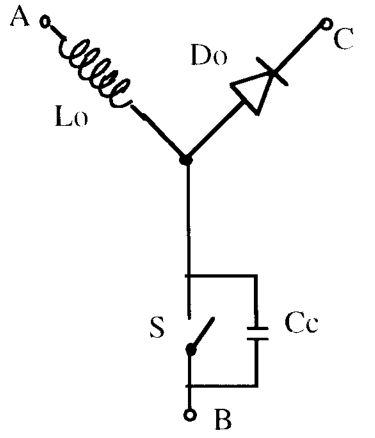
Can parasitic elements be used ?

1. 'Turn on'



- L_S relatively large inductance to lower $\frac{dI}{dt}$ and hence diode peak reverse current and trapped energy
- Need a mechanism for recovering trapped energy
  - Impractical to obtain by stray inductance such as interconnections
  - Can be part of transformer or flyback coupled inductor

2. 'Turn off'



- C_c significant inductance to lower $\frac{dV}{dt}$
- C_c is mainly needed while voltage on switch is low

👁 Switch output capacitance would be helpful and sometime sufficient

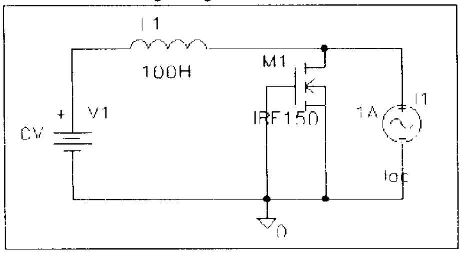
2. 'Turn off' - cont.

👁 If internal capacitance is used:

- fast gate 'turn off'
- high sink current
- low source inductance "fast switch"

👁 Look up new MOSFET generations (APT V)

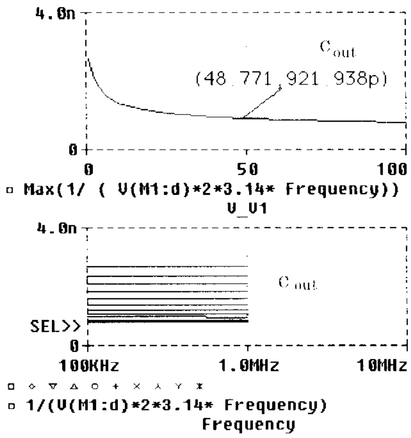
Output Capacitance of MOSFETs
A study by Simulation



- Nested AC analysis
- V_1 is swept from 0V to 100V
- I_{ac} is set to 1A (not to worry, used after linearization)
- Capacitance is calculated from:

$$C_{out} = \frac{1}{2\,\pi\ \text{frequency}\ v_{ds}}$$

'frequency' -> PSPICE variable

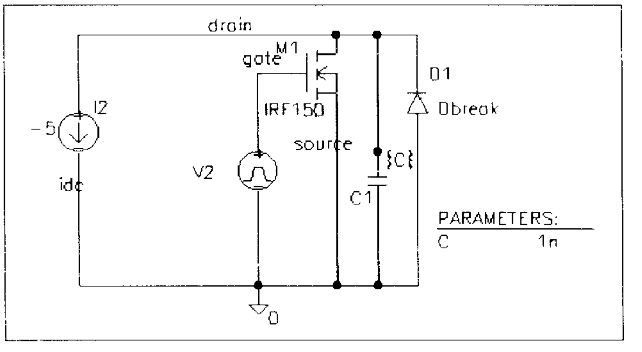


Upper trace: C_{out} as a function of V_{DS}

Lower trace: C_{out} as a function of frequency for various V_{DS} values.

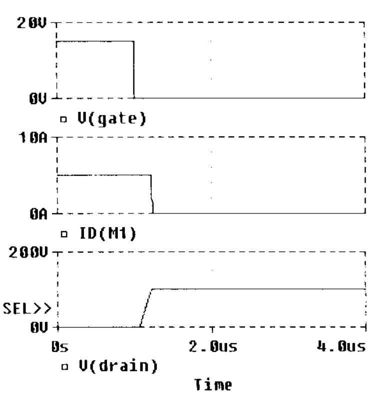
- About 1 nF at 50V
- For significant improvement added snubber capacitance > 1nF

Simulating Snubbing Effects 'turn-off'

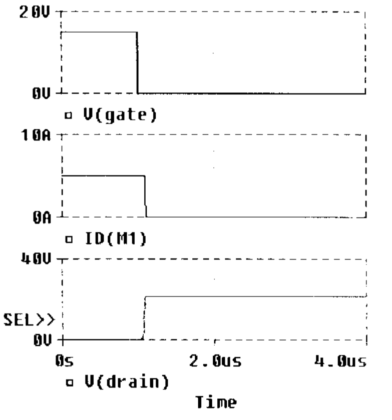


- Transient analysis
- C is swept 0nF - 10nF

C=0



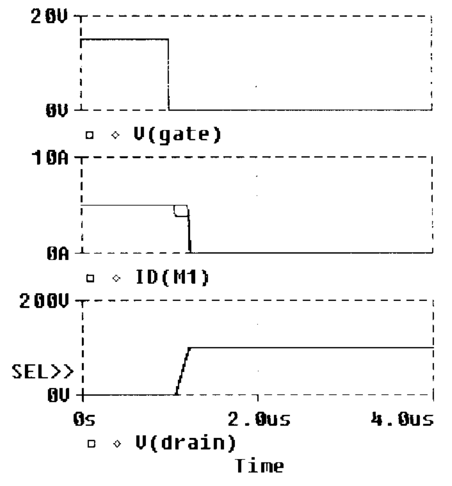
Drain clamped to 100V



Drain clamped to 20V

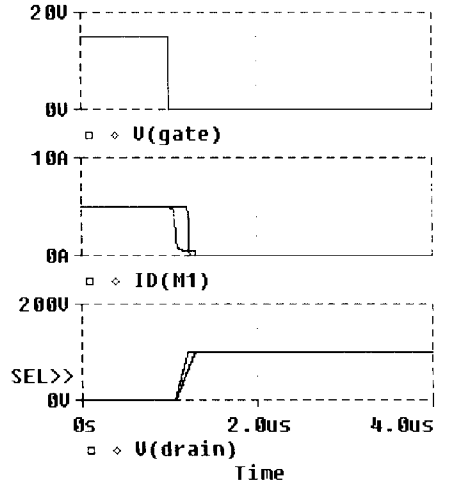
- Drain current reflects charging of internal capacitor
- Part of energy stored as $\frac{C\ V^2}{2}$ --> losses at 'turn-on' !
- Not all of $\{I_D{*}V_{DS}\}$ at 'turn off' duration is lost to heat

C = 0 & 1nF



- Small part of current channeled to snubbing capacitor

C = 0 &10nF



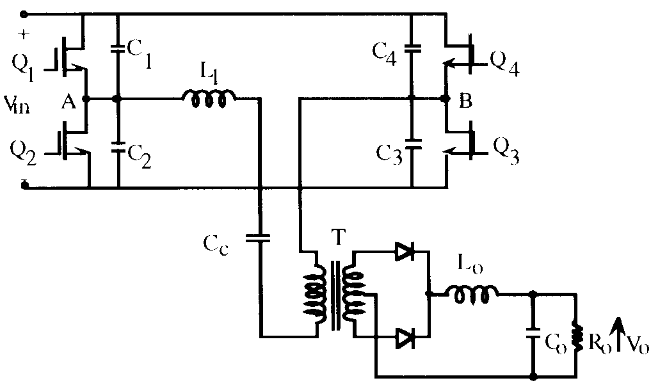
- Current channeled to 10nF capacitor
- Internal capacitor still charges to the same voltage !

Chapter 7

RECTIFIER DIODES LOSSLESS SNUBBERS

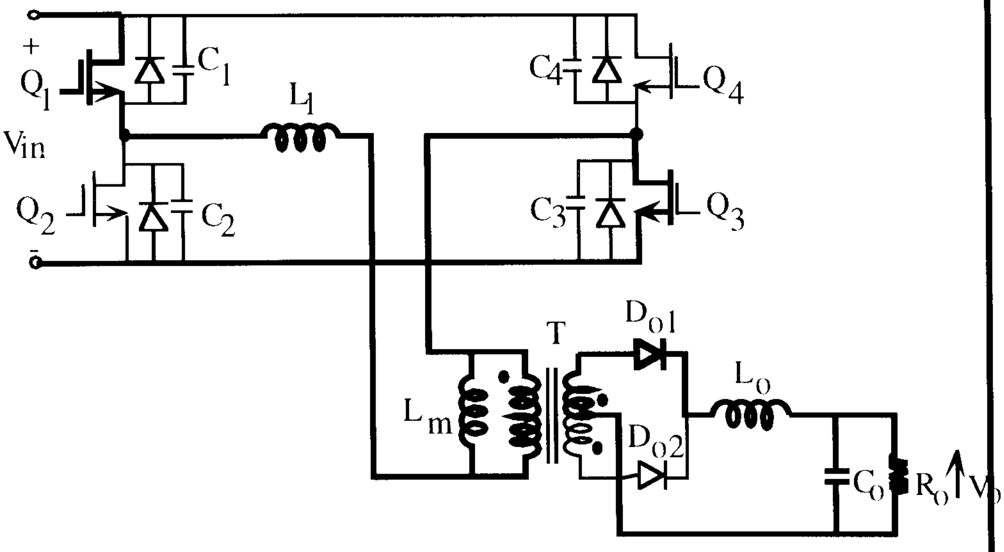
PHASE SHIFTED PWM (PSPWM) [18, 19, 27]

ZVS of main switches

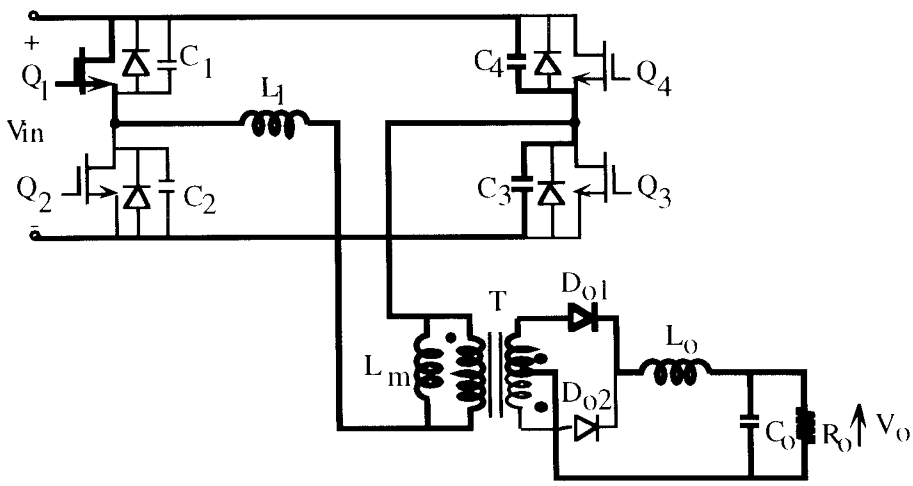


Basic Phase Shifted PWM Stage

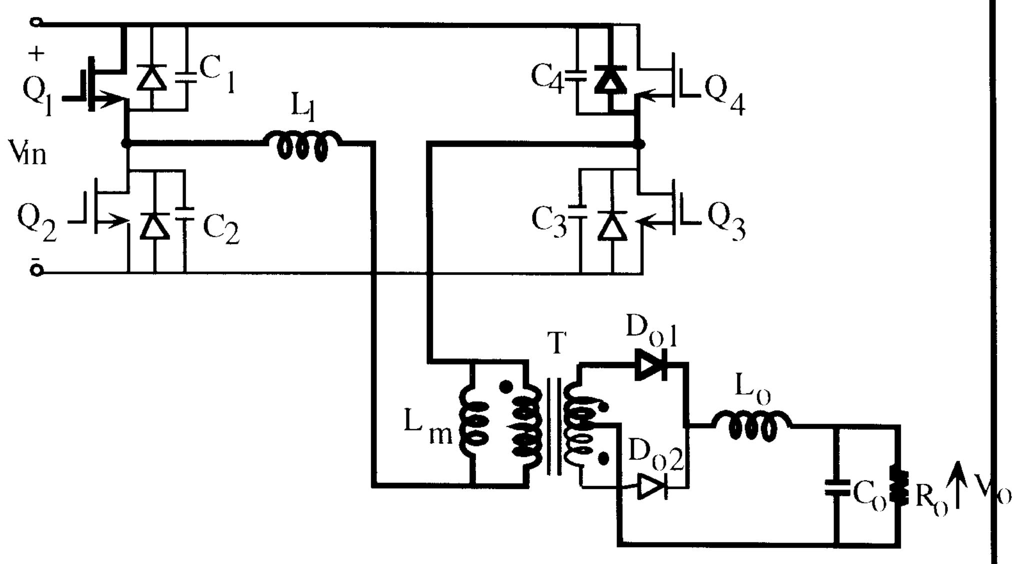
- Leading and lagging legs (bridge non-symmetrical)
- Coupling capacitor C_c needed to block current
    - Asymmetry of control (+noise !)
    - Can be eliminated by applying peak current mode control



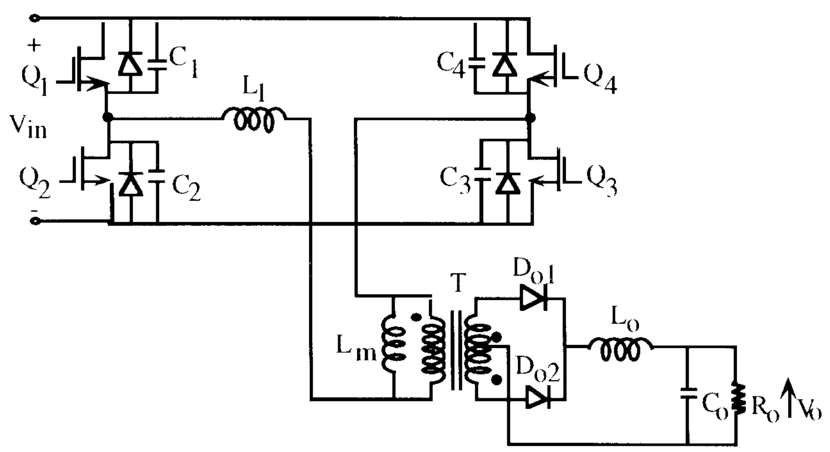
Stage 1 - Powering



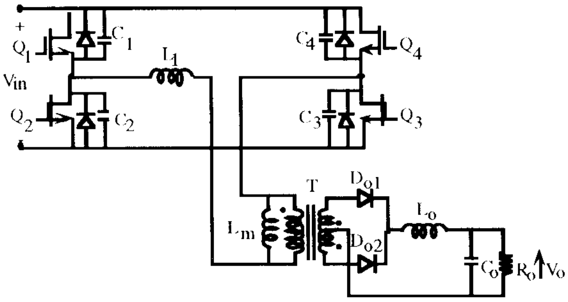
Stage 2 - Power to t_{off} transition - capacitor charge



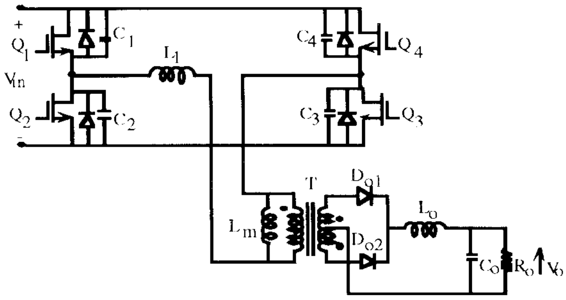
Stage 3 - Power to t_{off} transition - Diode catch



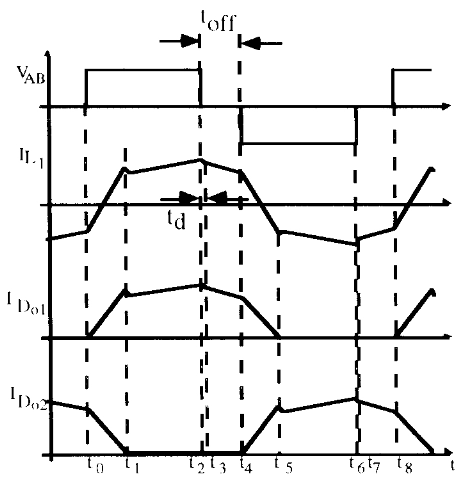
Stage 4 - t_{off} (circulating current)



Stage 5 - t_{off} to Power Transition



Back to Power Mode



Waveforms

✎ Two rectifier diodes are conducting during t_{off}

PSPWM DESIGN OVERVIEW

Basic Conventional Phase Shift PWM design equations

$$\frac{I_{off}^2 L_r}{2} \geq \frac{(C_A + C_B)V_{BUS}^2}{2}$$

C_A+C_B = total leg capacitance

or:

$$I_{off}\sqrt{\frac{L_r}{C_A + C_B}} > V_{BUS} \quad ==> \quad I_{off} Z_r > V_{BUS}$$

where the characteristic impedance is defined as usual:

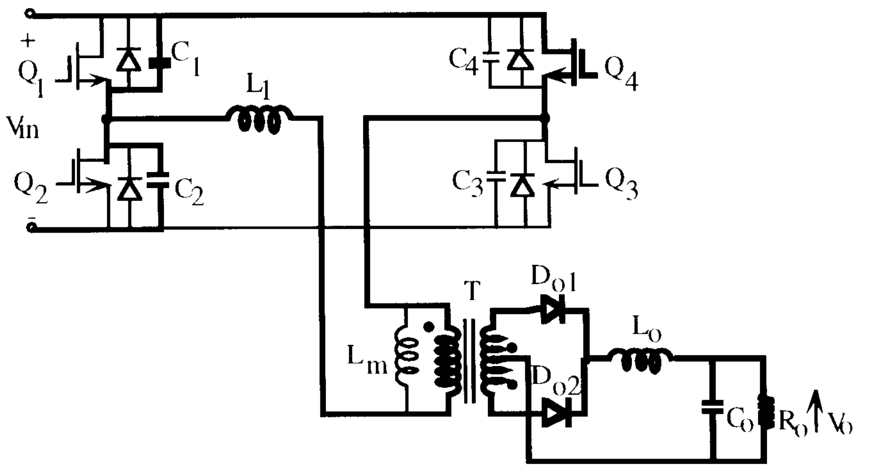
$$Z_r = \sqrt{\frac{L_r}{C_A + C_B}}$$

The transition time (need to be larger than 1/4 of the resonant period

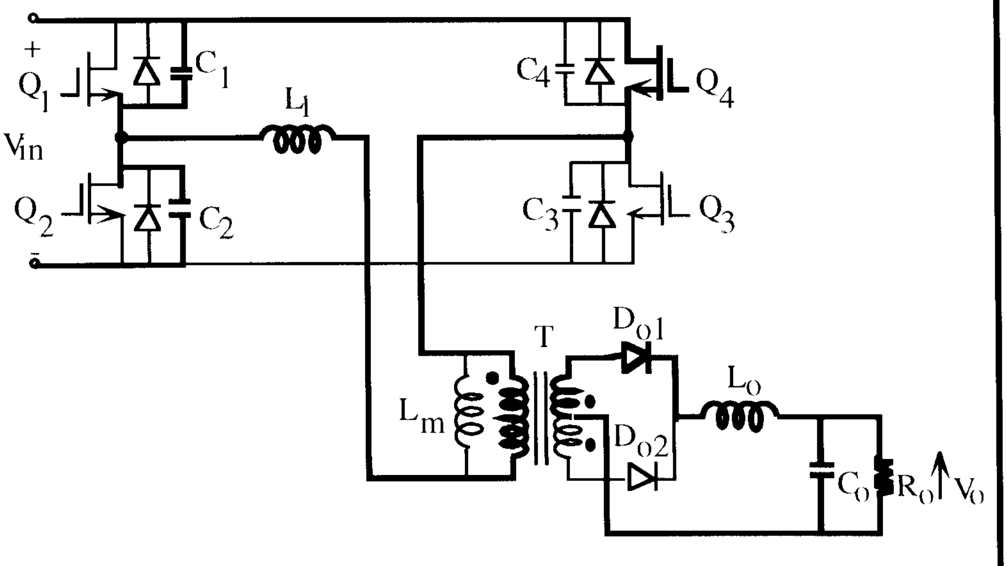
$$t_d \geq \frac{2\pi\sqrt{L_r(C_A + C_B)}}{4}$$

In the conventional scheme, L_1 need to be relatively large to store sufficient energy at low load current . This can be somewhat improved by making the primary current larger (lower L_m transformers magnetization inductance) .

If the short circuit (two diodes conducting) during t_{off} is avoided (only one diode conducts) L_1 can be small since the energy comes from the secondary



Stage 5 - t_{off} to Power Transition - two diodes conducting

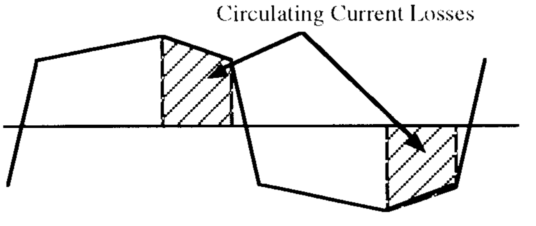


Stage 5 - t_{off} to Power Transition - one diode conducting

If only one diode conducting ==> L_1 id not needed

RESIDUAL ISSUES IN PSPWM

1. Circulating Current

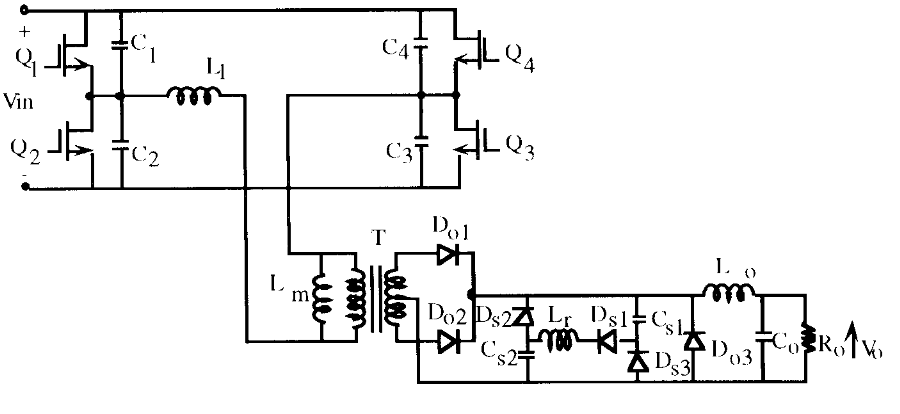


Primary current

- Thermal design of power stage for duty cycle D ==>1
  =>The circulating current at D < 1 not a major problem

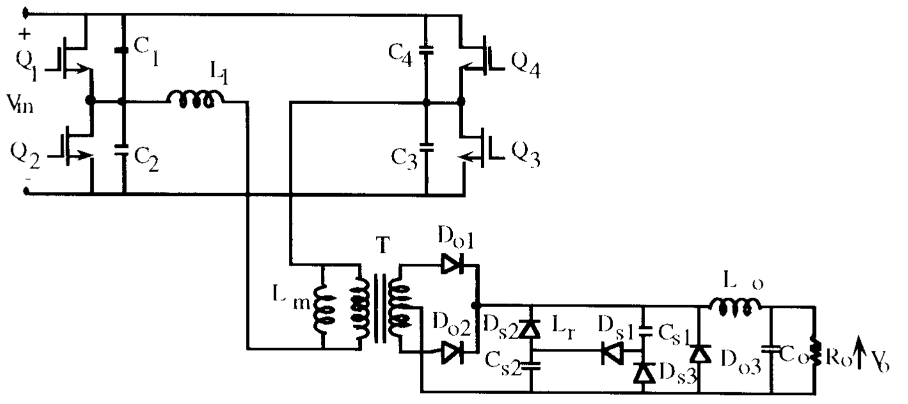
- Possible Solution

Apply the 'One Way' capacitor to block primary current at t_{off}



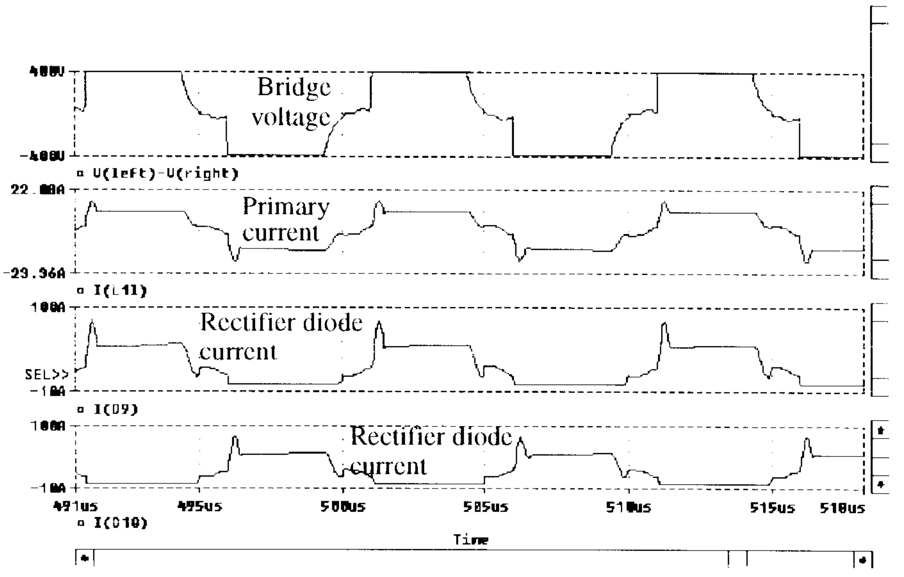
Lossless output snubber SNB16 [18]

- L_r and C_s form a resonant circuit



Modified lossless output snubber SNB17 [19]

Resonant inductor can be moved to primary

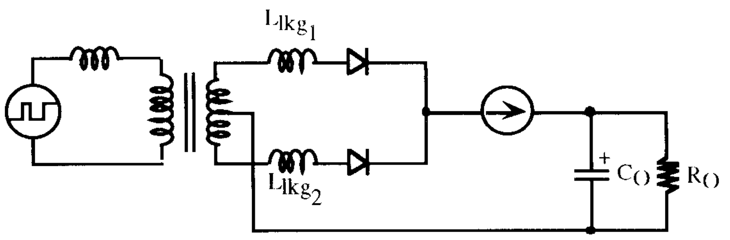


Basic waveforms of SNB17 (Simulation)

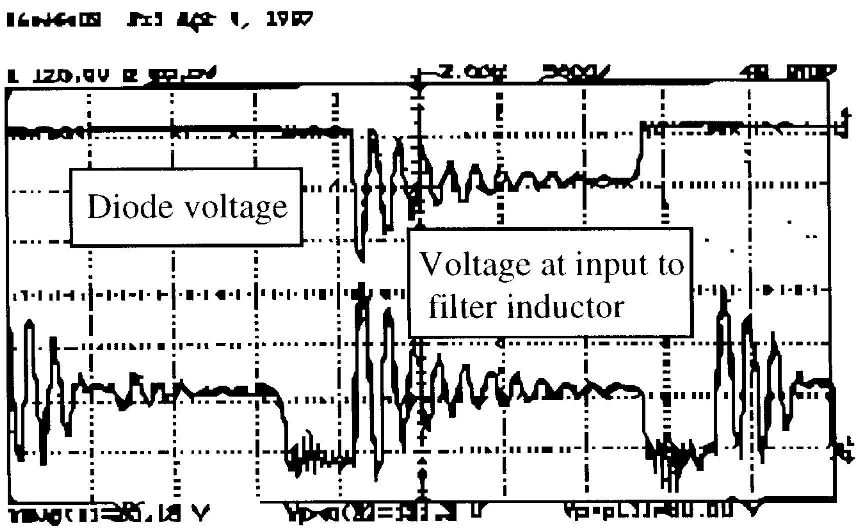
- Suitable for IGBT (zero current at t_{off} combats current tail)

RESIDUAL ISSUES IN PSPWM

2. Rectifier's diodes reverse recovery



Leakage inductance at secondary

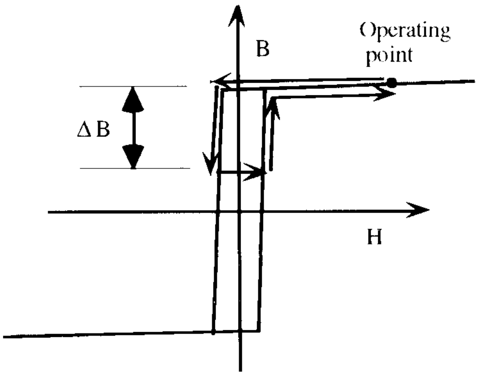


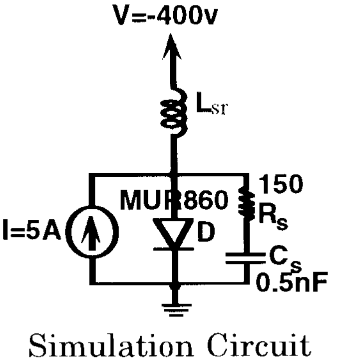
Experimental waveforms

- Greatly influenced by transformer leakage

RESIDUAL ISSUES IN PSPWM

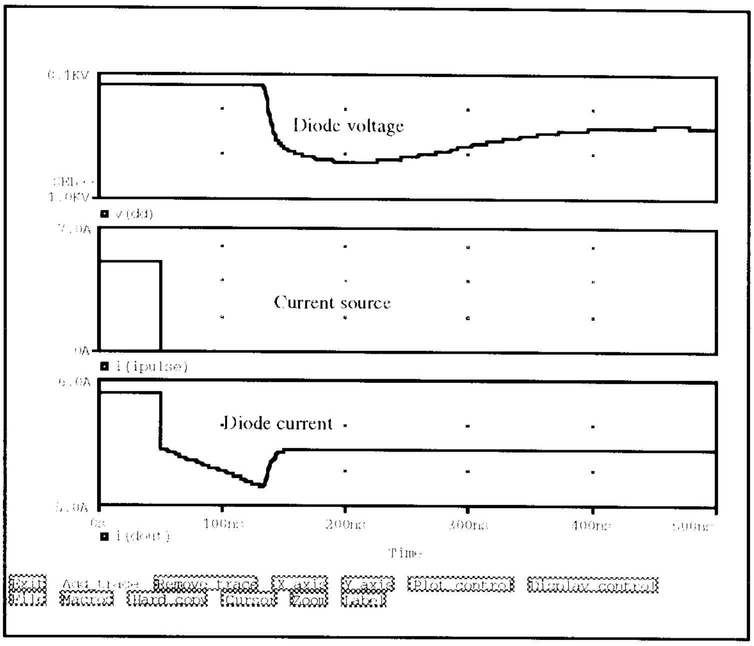
2. Rectifiers' diodes reverse recovery
   - Possible Remedy -> Saturable reactor (SR) [7, 8, 9, 14, 43]





Simulation Circuit

- Using fast diode
- Current source establishes the forward current and abrupt change
- L_{sr} emulate "large" back inductance



Simulation waveforms (L_{sr}= 50μH)

RESIDUAL ISSUES IN PSPW

2. Rectifiers' diodes reverse recovery

- Design equations overview

Magnetic flux swing (one sided)

$$\Delta B = \frac{VT}{2\ n\ A_c}$$

VT = volt-second across SR

A_c= core effective cross section area

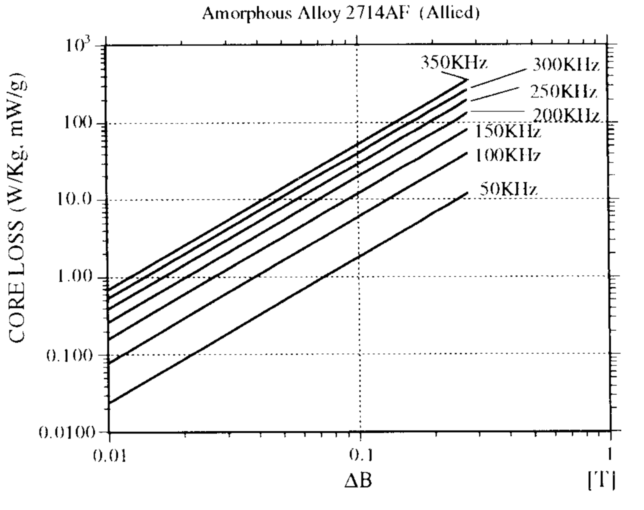
n= number of turns

For Amorphous Alloy 2714AF (Allied):

Core Losses : Pcore (W/Kg, mW/gr)) = 10^{-6} $(f_s)^{1.73}$ $(\Delta B)^{1.88}$

B = Magnetic flux (T) ; f_s = Switching frequency (Hz)

$\Delta B=(Pcore/(10^{-6}\ (f_s)^{1.73}))^{(1/1.88)}$

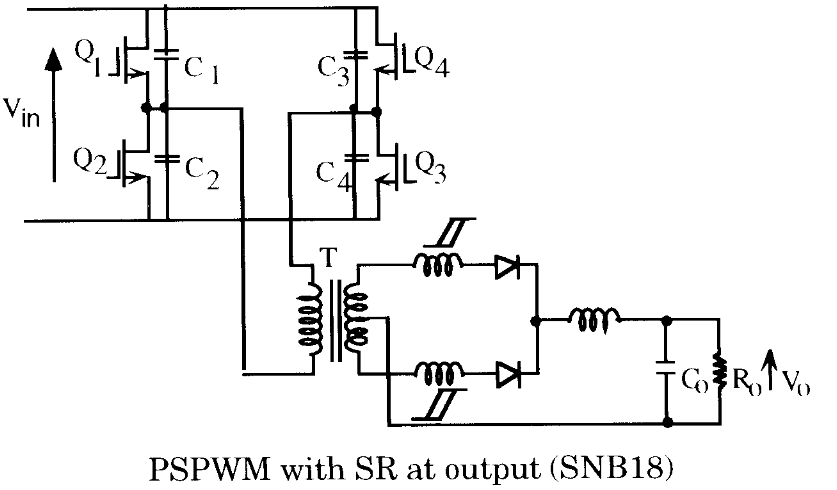


Extrapolated from Allied's data (originally given up to 200KHz)

RESIDUAL ISSUES IN PSPWM [42, 43]

2. Rectifiers' diodes reverse recovery

- Implementation in PSPWM

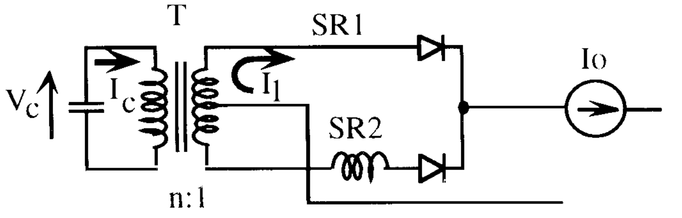


PSPWM with SR at output (SNB18)

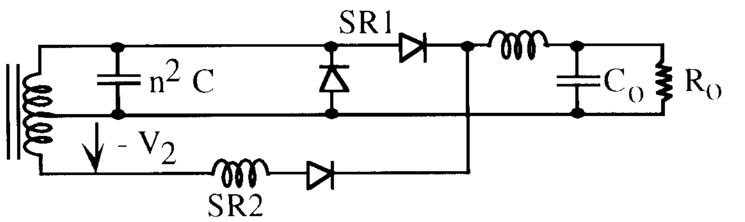
RESIDUAL ISSUES IN PSPWM

2. Rectifiers' diodes reverse recovery

- Implementation in PSPWM



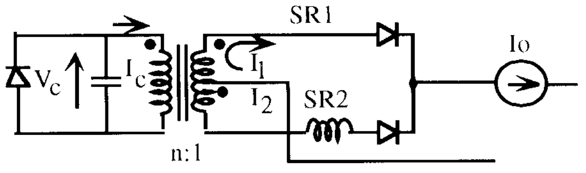
Situation at first commutation instance (power to t_{off})



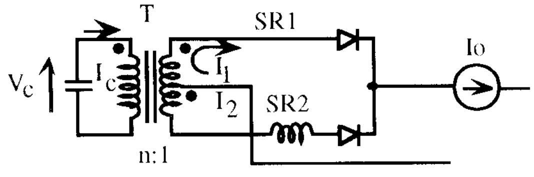
Equivalent circuit during t_{off}

RESIDUAL ISSUES IN PSPWM

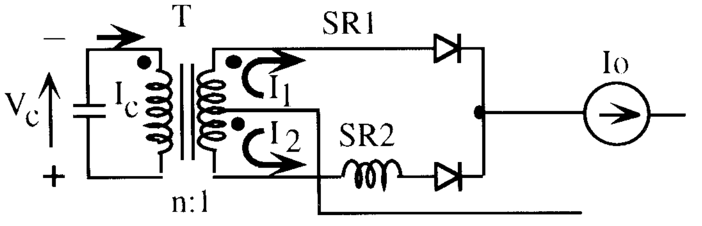
- Theoretical considerations



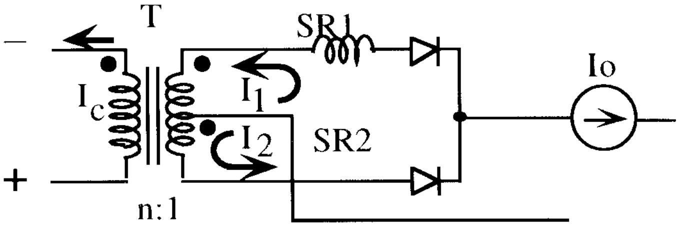
t_{off}



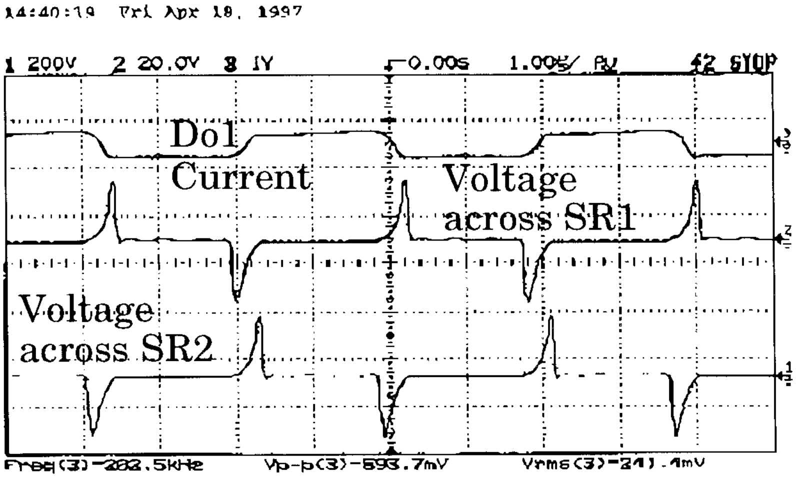
t_{off} release



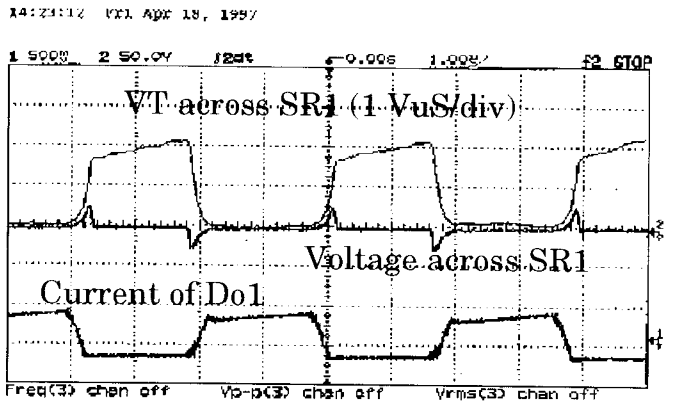
t_{off} to Power transition



Transition end (SR1 blocks reverse recovery current)



Current through rectifier leg and voltage across SR



Volt Second of SR

Design Guidelines

1. Choose SR to keep core losses low say : 50 mW/gr
2. Estimate VT

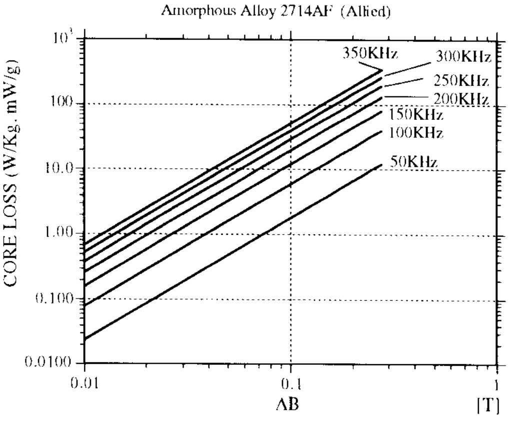
Example: Diode reverse voltage 100V

$t_{rr} \approx 100$ nSec => $V\,T \approx 10$ VμS

For Amorphous Alloy 2714AF (Allied):

$$\text{Pcore (W/Kg, mW/gr))} = 10^{-6}\ f^{1.73}\ \Delta B^{1.88}$$

Assume Pcore = 50 mW/gr; f = 270KHz



From loss equation => $\Delta B = 0.125T$

From $\Delta B = \dfrac{V\,T}{2\,n\,A_c}$ => $A_c = 0.4\ cm^2$

Using MP1906P-4AF (Allied):

OD = 21mm; ID = 10mm; $A_c = 0.16\ cm^2$; n=1; 6.1 gr

- ☞ Need at least 3 units per leg
- ☞ Effective frequency is much higher than switching frequency
- ☞ Being a small body (small surface area), might get very hot. 4 units per leg were used in experimental converter.

Experimental PSPWM Converter

Transformer: Payton 3000W T250-12-4C

Nominal Operating frequency = 350KHz

Input Voltage = 360 -390 Volt

Max. VT = 864 VμS

Primary to half secondary 6:1

Primary Max rms current: 11 Amp

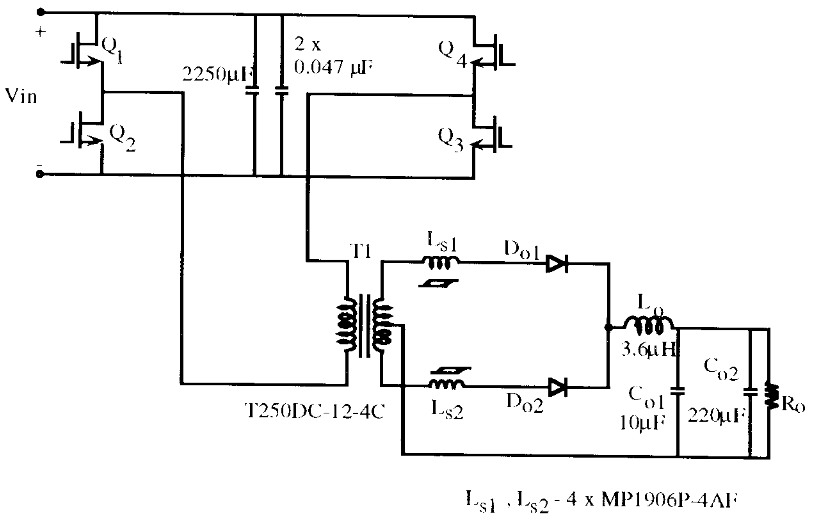
Dielectric strength (primary to secondary) 2500 Vrms

Dielectric strength (secondary to core) 500 Vdc

Estimated power loss 65W (@ 60^0C base plate)

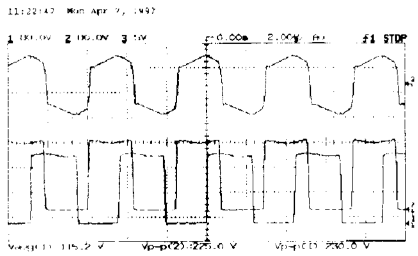
Estimated hot spot 110^0C

Mechanical dimensions 88*65*30(h) mm

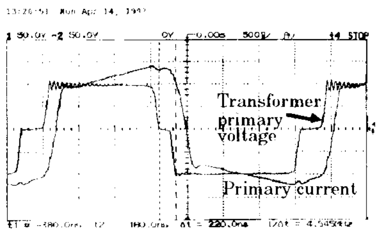


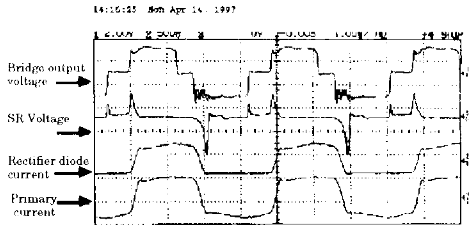
Experimental power stage

Experimental Results

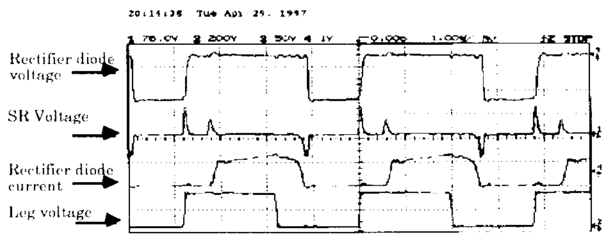


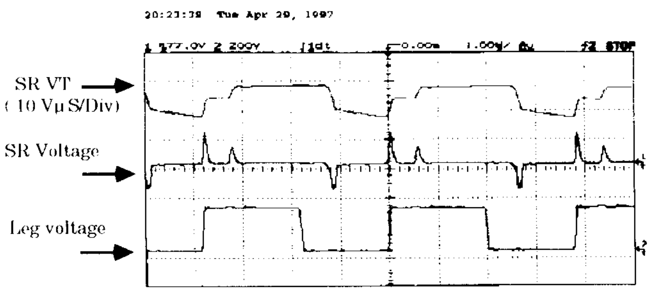
Primary current (upper) and leg voltages (lower)



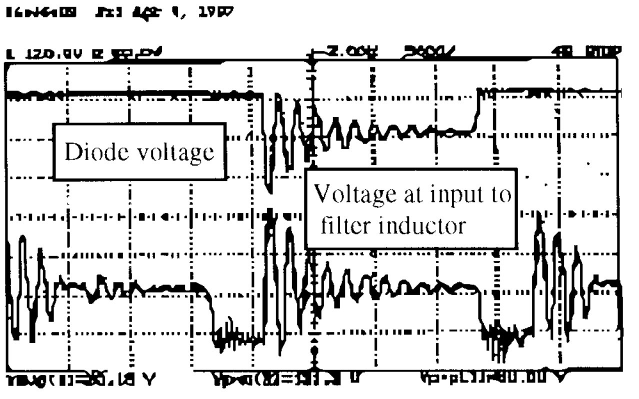


In circuit SR Performance`

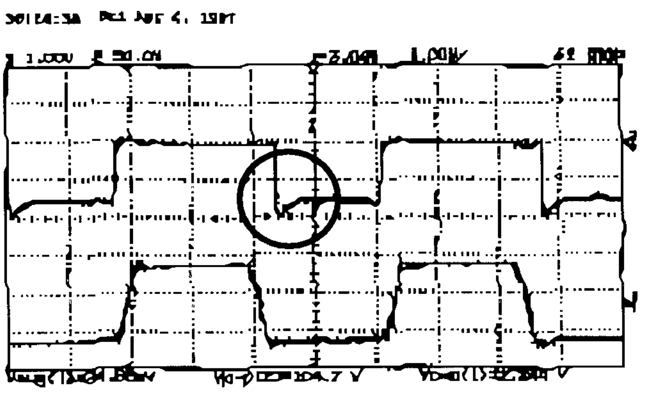




Experimental Results



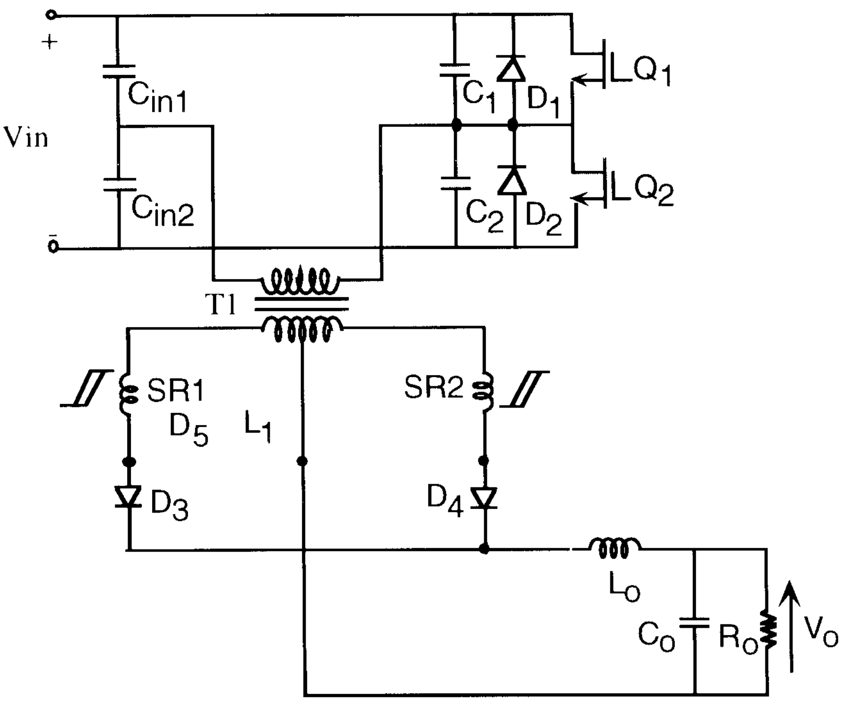
Experimental waveforms (no SR snubber)



Rectifier diode voltage (upper trace) and current (lower) with amorphous core

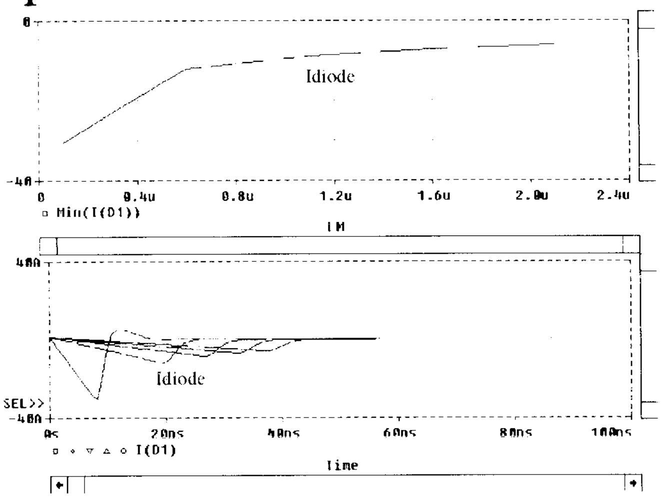
- Some overshoot due to energy stored in SR

Improved magnetic snubber for rectifier diodes
in DC/DC converters



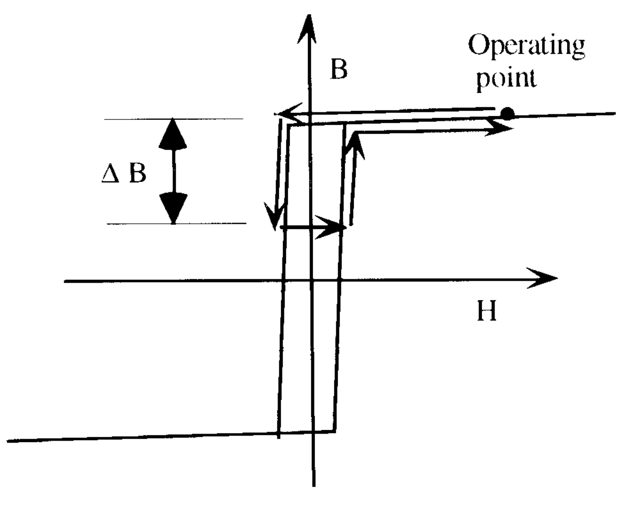
- Basic solution [42,43]

Dependence on L - Simulation



Diode reverse recovery switching waveforms

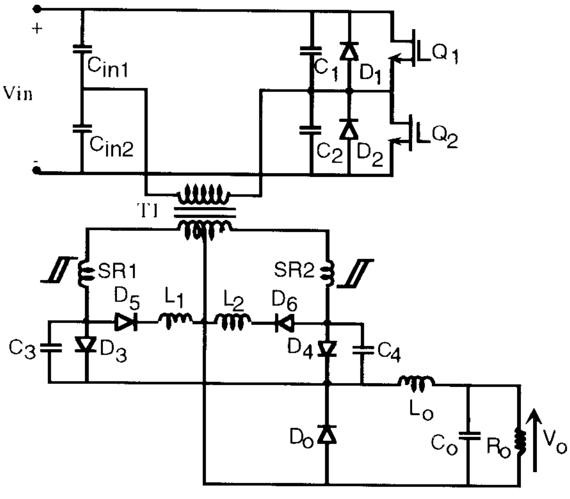
- Actual reverse recovery time is a function of series inductance



- Long reverse recovery time will increase core losses

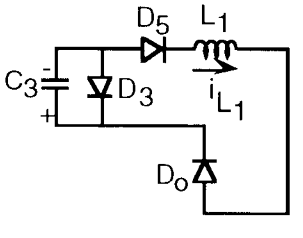
APEC

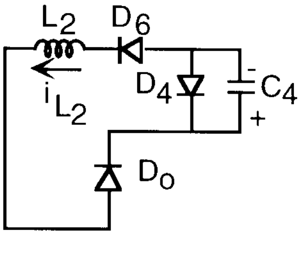
Improved magnetic snubber for rectifier diodes in DC/DC converters [44]



Added elements: C_3, C_4, D_o, L_1 D_5 & L_2 D_6

-> Extra resonant circuit to sweep out stored charge



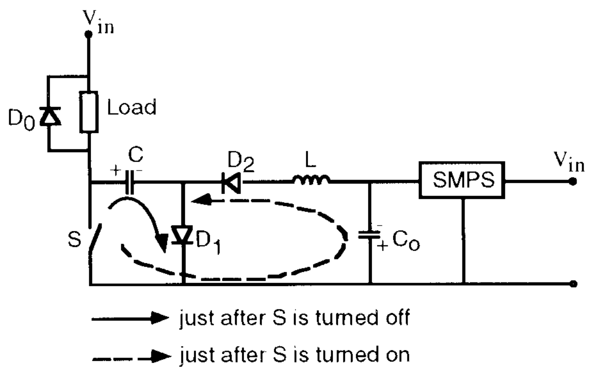


The stored charge of the diode D_3 (D_4) is pulled out rapidly ; therefore (t_{rr}) D3 and (t_{rr}) D4 desrease and hence core losses in SR1 and SR2 also reduce.

Chapter 8

COMBINING SNUBBER AND POWER SUPPLY FUNCTIONS

Turn-off snubber with energy recovery back to the input bus or into a local power supply (SNB20) [39]



C = Snubber capacitor

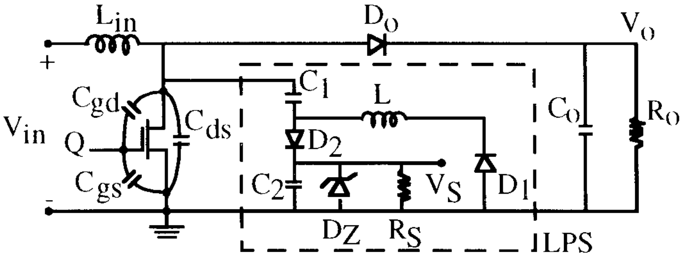
C_0 = Filter capacitor

Negative polarity of voltage across C_0 !

- The main purpose of the device: turn-off snubber
- Additional appointment: local power supply

Watch for power level (in PS applications), might be too high

A local power supply with turn off snubber features [47]



LPS - local power supply

R_S - load resistance

$C_2 >> C_1$; D_Z - Zener diode

- Positive polarity of output voltage V_S
- The main purpose of the device: A local power supply.

Secondary objective: turn-off snubber

A local power supply with turn off snubber features

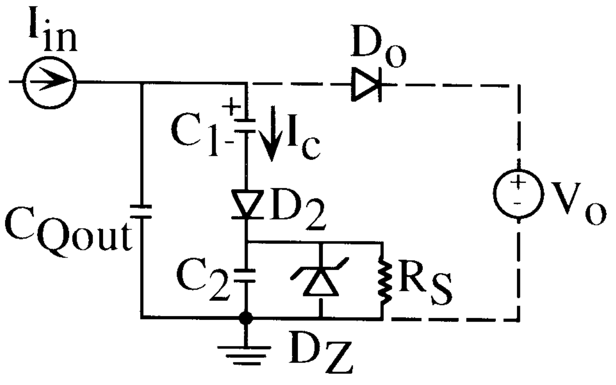
Main assumptions of analysis

1. Ideal diodes and ideal transistor. Parasitic capacitences of MOSFET are taken into account. It is assumed that these capacitences are linear.

2. Lin, C_0 and C_2 are infinitely high

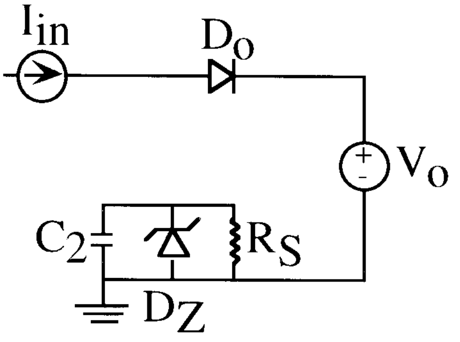
APEC

Equivalent circuits



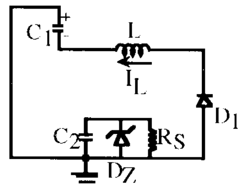
- t_0-t_1: energy injection into LPS

Equivalent circuits

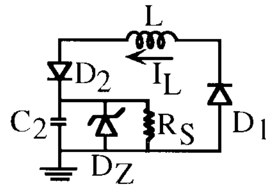


- t_1-t_2: no interconnection between the processes in the LPS and in the converter

Equivalent circuits

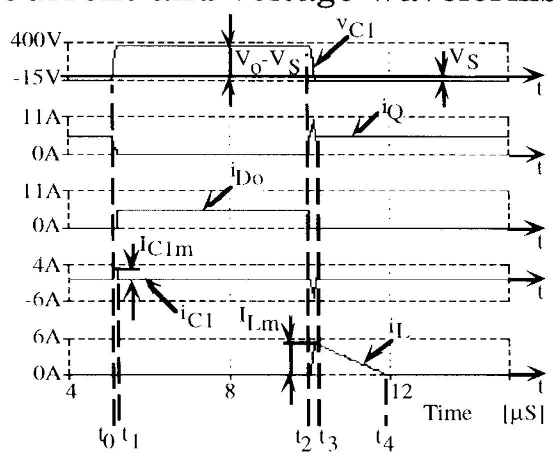


t_2-t_3: energy transfer from the capacitor C_1 to the inductor L



- t_3-t_4: energy transfer from the inductor L to the C_2 - D_2 - R_s circuit

Current and voltage waveforms



Proposed Local Power Supply (LPS) connected in a boost converter

Main relationships

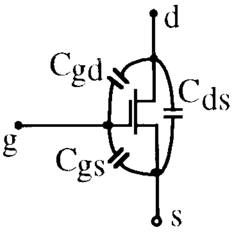
Energy injected to LPS during one switching period

$$E = \frac{C_1 V_o^2}{2} = V_s(I_s + I_Z)T_s$$

$$I_s + I_Z = f_s C_1 \frac{V_o^2}{2V_s}$$

$$I_{s\ max} = f_s C_1 \frac{V_o^2}{2V_s} \quad \text{(when } I_Z=0\text{)}$$

LPS loaded by the MOSFET's controller driver



The average input current to the gate of the MOSFET

$$I_{g\ av} = C_{Qin} V_{gs} f_s = k I_{s\ max}$$

k= fraction of I_s used to power the gate (k<1)

The required value of C_1

$$C_1 = \frac{2}{k}\frac{V_{gs}}{V_o}\left[C_{gd}\left(1 + \frac{V_{gs}}{V_o}\right) + C_{gs}\frac{V_{gs}}{V_o}\right]$$

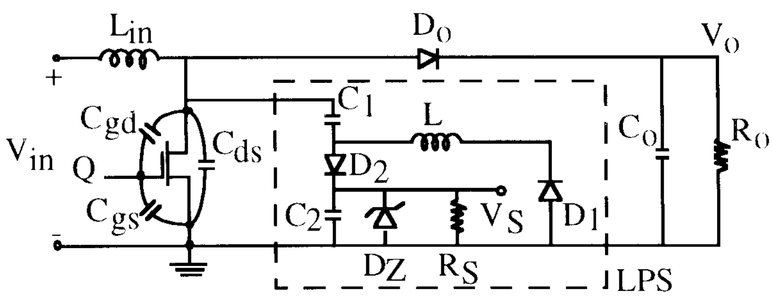
If $V_{gs} << V_o$

$$C_1 \approx \frac{2}{k}\frac{V_{gs}}{V_o}C_{gd}$$

C_1 is of the same order of magnitude as the parasitic capacitences of the transistor!

If additional power is required (e. g. DC fans) C_1 will be larger.

Experimental boost converter with proposed LPS



Q=IRFP460; D_0=MUR460; D_1=1N5819; D_2=MUR160

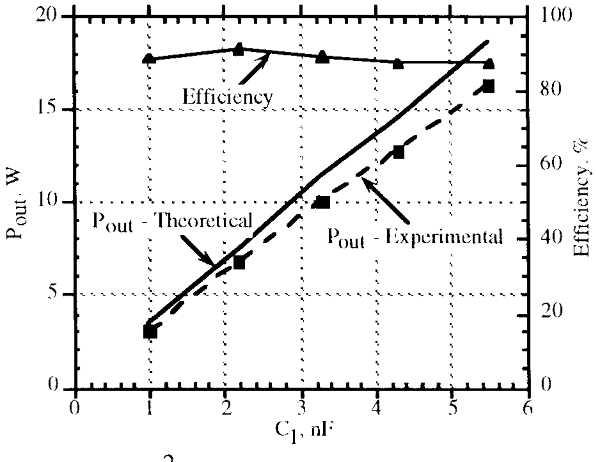
Power stage:

f_s=100kHz; L_{in}=1mH; L=24.2μH; C_0=1mF; V_o = 260V; P_0=85W

Snubber:

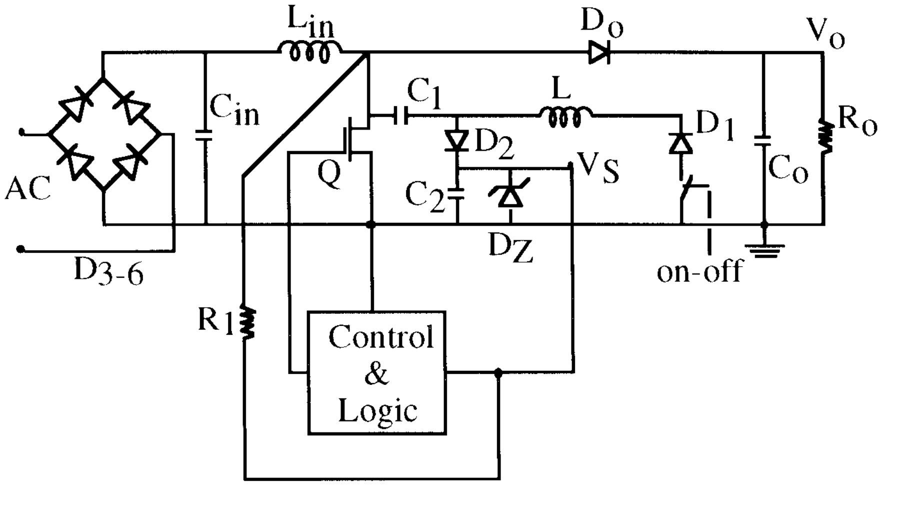
V_s=15V; C_2=100μF; R_s=10-64Ω ; C_1=1.0-5.5nF

Output power and efficiency as functions of the charge pump capacitor C_1

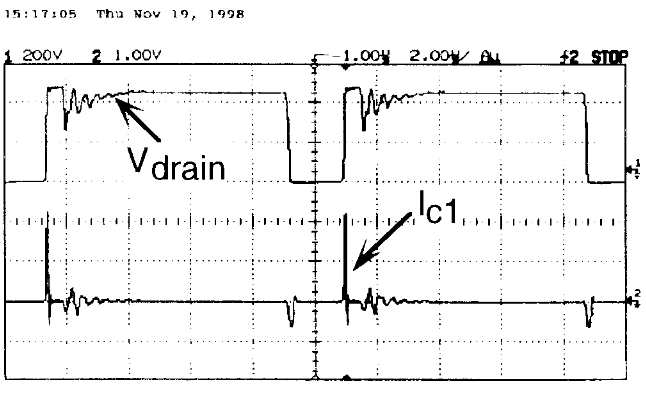


$$P_{out(theor)} = f_s E = \frac{f_s C_1 V_o^2}{2}$$

Application of proposed local power supply in a boost APFC



Experimental waveforms of the LPS operating in a soft switched APFC circuit



$C_1 \approx 220pF$, $V_{in}=220V_{rms}$, Vo=380V, Vs=12.4V, Rs≈170Ω

CONCLUSIONS

- Passive lossless snubbers can improve performance of switch mode systems:
  - Controlled $\frac{di}{dt}$ and $\frac{dv}{dt}$
  - Increased efficiency at high switching frequency
  - Reduction of voltage and current spikes
  - Relatively low cost
  - High reliability
- The magnitude of the reverse recovery current of the diode may affect performance of the snubber
- Check points to watch:
  - Extra stresses
  - Duty cycle limitation
  - Current limitation

☞ Snubber/LPS combination could be useful in some applications

Methods to increase the snubbing capacitor without increasing LPS power - are now under investigation

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