Lecture 7: MOSFET, IGBT, and Switching Loss

ECE 481: Power Electronics

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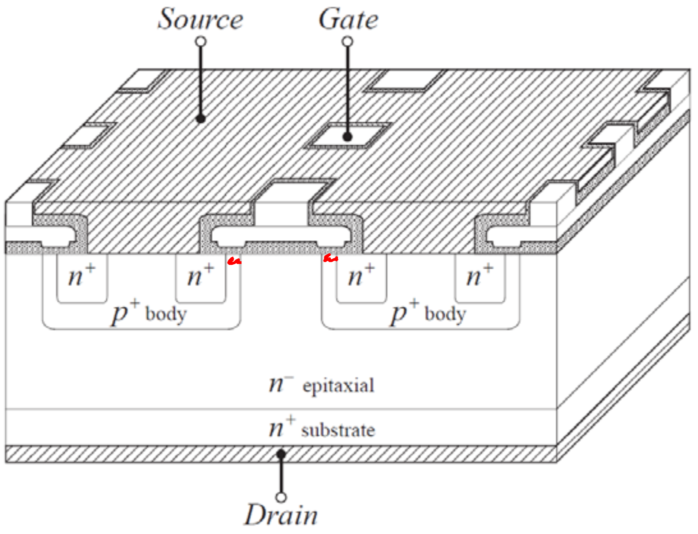
Fall 2013

Announcements

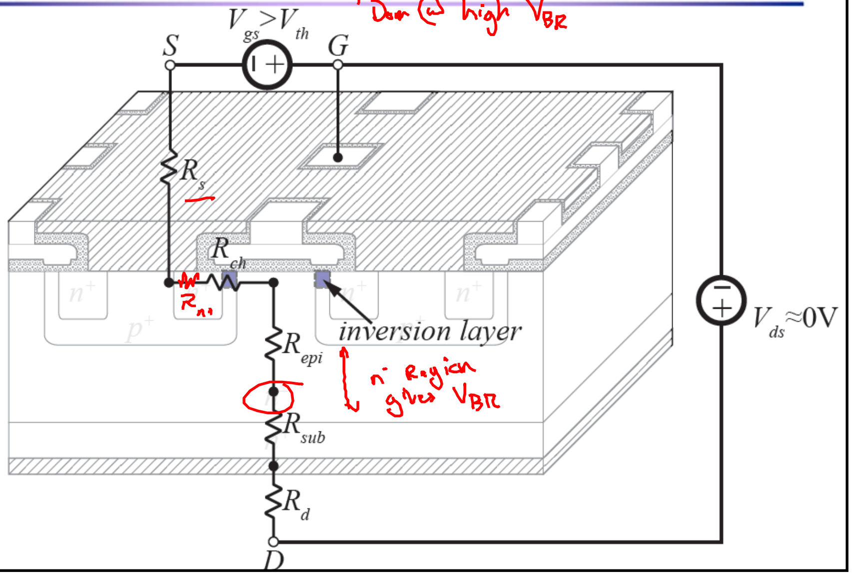
- Homework #3 posted this afternoon
  - Plots in homeworks: label all salient features
  - Assignment clarity: Box answers, staple sheets, include course number
- Course E-mail list
- No office hours tomorrow
- No class next week
- Midterm Exam 1 handed out on 9/26

MOSFET On-State

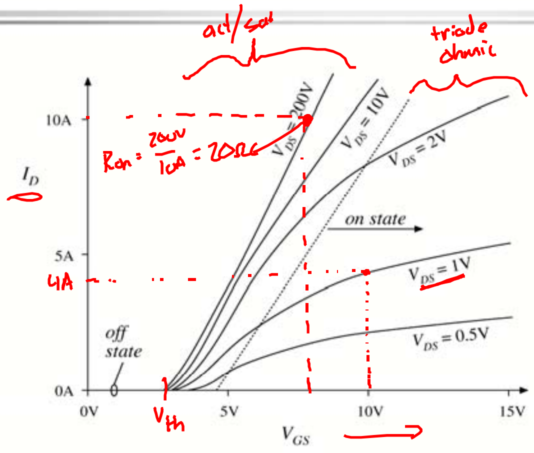


$$R_{on} = (R_s + R_{ch} + R_{n^+} + \underline{\underline{R_{epi}}} + R_{sub} + R_d)$$

↑ Dom @ high V_{BR}



Typical MOSFET characteristics



$R_{on} = \frac{1V}{4A} = 250\,m\Omega$

- Off state: $V_{GS} < V_{th}$
- On state: $V_{GS} >> V_{th}$
- MOSFET can conduct peak currents well in excess of average current rating —characteristics are unchanged
- on-resistance has positive temperature coefficient, hence easy to parallel

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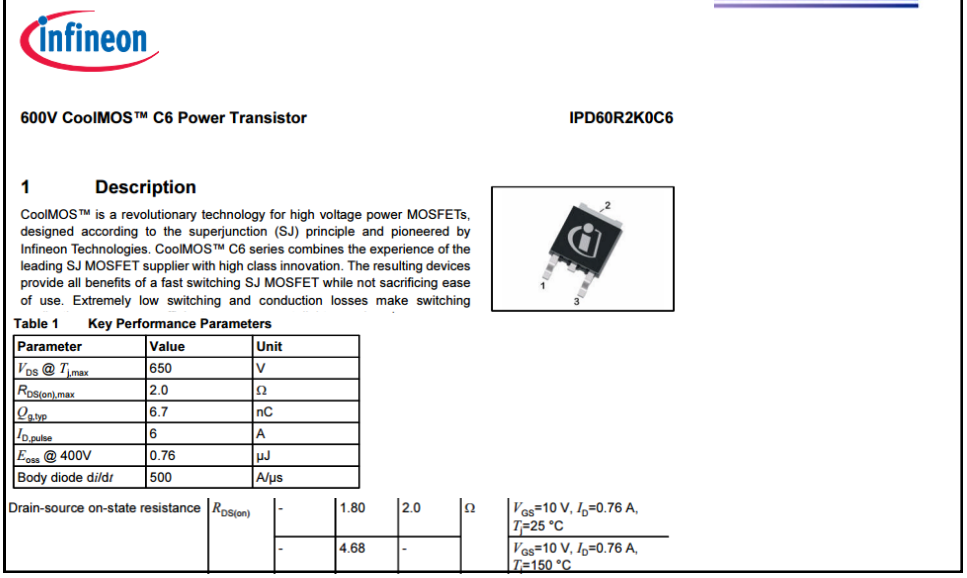
MOSFET Datasheet



600V CoolMOS™ C6 Power Transistor **IPD60R2K0C6**

1 Description

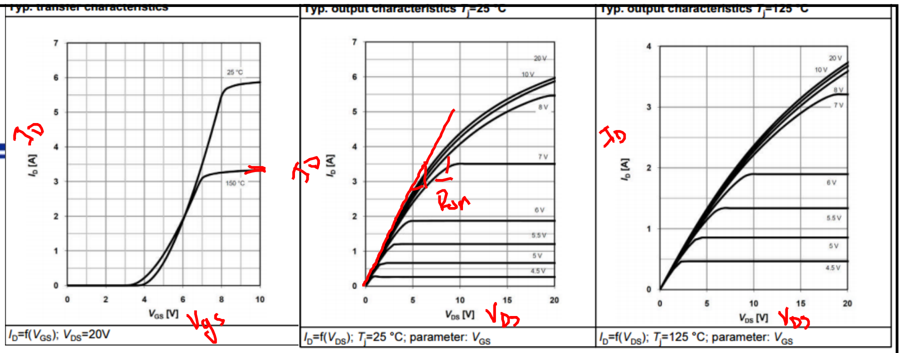
CoolMOS™ is a revolutionary technology for high voltage power MOSFETs, designed according to the superjunction (SJ) principle and pioneered by Infineon Technologies. CoolMOS™ C6 series combines the experience of the leading SJ MOSFET supplier with high class innovation. The resulting devices provide all benefits of a fast switching SJ MOSFET while not sacrificing ease of use. Extremely low switching and conduction losses make switching

Table 1 Key Performance Parameters

Parameter	Value	Unit
V_{DS} @ $T_{j,max}$	650	V
$R_{DS(on),max}$	2.0	Ω
$Q_{g,typ}$	6.7	nC
$I_{D,pulse}$	6	A
E_{oss} @ 400V	0.76	µJ
Body diode d*i*/d*t*	500	A/µs

Drain-source on-state resistance	$R_{DS(on)}$	-	1.80	2.0	Ω	V_{GS}=10 V, I_D=0.76 A, T_j=25 °C
		-	4.68	-		V_{GS}=10 V, I_D=0.76 A, T_j=150 °C

Typ. transfer characteristics | Typ. output characteristics T_j=25 °C | Typ. output characteristics T_j=125 °C



I_D=f(V_{GS}); V_{DS}=20V

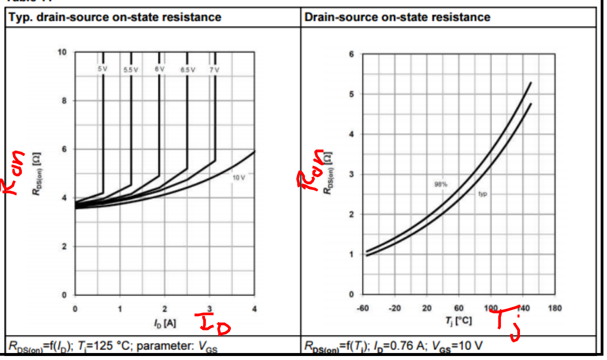
I_D=f(V_{DS}); T_j=25 °C; parameter: V_{GS}

I_D=f(V_{DS}); T_j=125 °C; parameter: V_{GS}

$R_{on} = \frac{V_{DS}}{I_D}$

Table 11

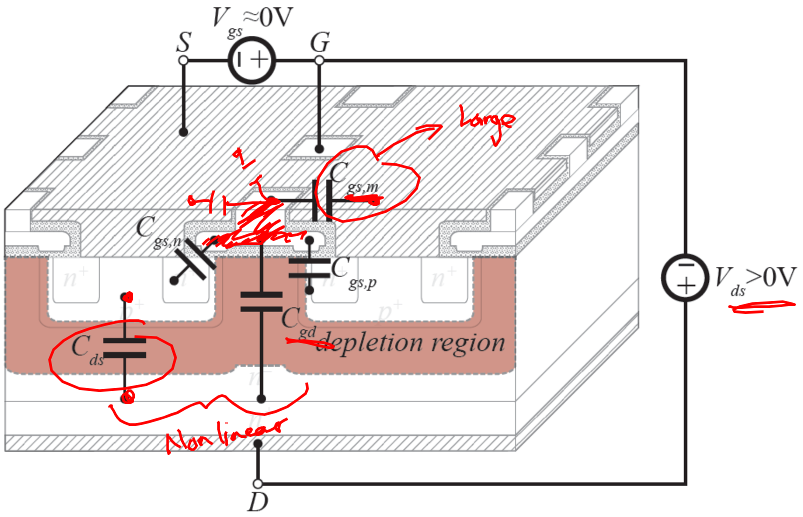
Typ. drain-source on-state resistance | Drain-source on-state resistance



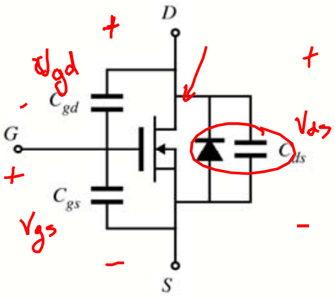
$R_{DS(on)}$=f(I_D); T_j=125 °C; parameter: V_{GS}

$R_{DS(on)}$=f(T_j); I_D=0.76 A; V_{GS}=10 V

MOSFET : OFF - State

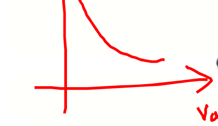


A simple MOSFET equivalent circuit



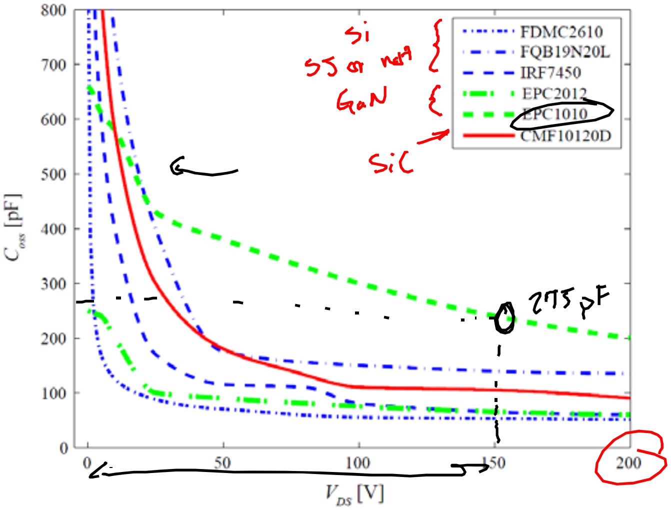
- C_{gs} : large, essentially constant
- C_{gd} : small, highly nonlinear
- C_{ds} : intermediate in value, highly nonlinear
- switching times determined by rate at which gate driver charges/discharges C_{gs} and C_{gd}

$C_{ds}(v_{ds})$



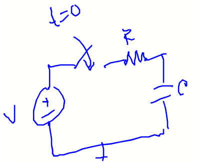
$$C_{ds}(v_{ds}) = \frac{C_0}{\sqrt{1+\frac{v_{ds}}{V_0}}}$$

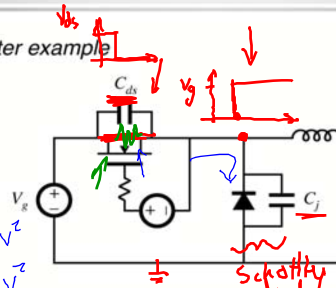
$$C_{ds}(v_{ds}) \approx C_0\sqrt{\frac{V_0}{v_{ds}}} = \frac{C_0'}{\sqrt{v_{ds}}}$$



Switching loss caused by semiconductor output capacitances

Buck converter example





Schottky diode (majority)

Before MOSFET turn-on

$V_{ds} = V_g$

$W_{cf} = E_c = \frac{1}{2} C_{ds} V_g^2$

$+$

$V < V_g$

$-$

at $t = \infty$ $E_c = \frac{1}{2} C V^2$

$E_R = \frac{1}{2} C V^2$

Energy lost during MOSFET turn-on transition (assuming linear capacitances):

$$W_C = \frac{1}{2} (C_{ds} + C_j) V_g^2$$

$\rightarrow P_c = W_c f_s$

$E_j = \frac{1}{2} C_j V_g^2$

$i_c = C \frac{dv}{dt}$

$P_R = i_c (V - V_c)$

$E_R = \int_0^\infty C (V - V_c) \frac{dv}{dt} dt$

$= \int_0^V C (V - V_c) dv$

$= \int_0^V C V dv - \int_0^V C V_c dv$

$= CV^2 - \frac{1}{2} C V^2 = \frac{1}{2} C V^2$

if you didn't believe

MOSFET nonlinear C_{ds}

Approximate dependence of incremental C_{ds} on v_{ds} :

$$C_{ds}(v_{ds}) \approx C_0 \sqrt{\frac{V_0}{v_{ds}}} = \frac{C_0'}{\sqrt{v_{ds}}}$$

Energy stored in C_{ds} at $v_{ds} = V_{DS}$:

$$W_{Cds} = \int v_{ds}\, i_C\, dt = \int_0^{V_{DS}} v_{ds}\, C_{ds}(v_{ds})\, dv_{ds}$$

$$W_{Cds} = \int_0^{V_{DS}} C_0'(v_{ds}) \sqrt{v_{ds}}\, dv_{ds} = \tfrac{2}{3} C_{ds}(V_{DS})\, V_{DS}^2$$

— same energy loss as linear capacitor having value $\frac{4}{3} C_{ds}(V_{DS})$

↓Q in C_{gs} ↓Q in C_{ds}

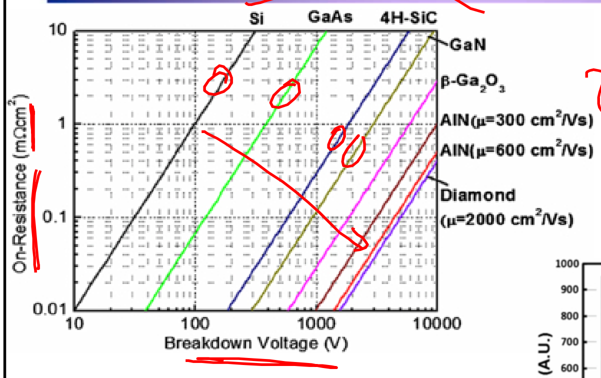
FOM → $Q_g \cdot r_{on}$ or $Q_{oss} r_{on}$

Characteristics of several commercial power MOSFETs

→ IRLHM620	20V	40A	0.0025Ω	52nC
GaN → EPC2015	40V	33A	0.004Ω	10.5nC

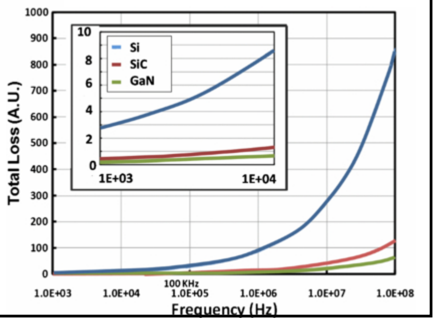
Part number	*Rated max voltage*	*Rated avg current*	R_{on}	Q_g *(typical)*
IRFZ48	60V	50A	0.018Ω	110nC
IRF510	100V	5.6A	0.54Ω	8.3nC
IRF540	100V	28A	0.077Ω	72nC
APT10M25BNR	100V	75A	0.025Ω	171nC
IRF740	400V	10A	0.55Ω	63nC
MTM15N40E	400V	15A	0.3Ω	110nC
APT5025BN	500V	23A	0.25Ω	83nC
APT1001RBNR	1000V	11A	1.0Ω	150nC
superjunction { IPW60R099CP	600V	31A	0.1Ω	60nC
IPW90R340C3	900V	15A	0.34Ω	93nC
GaN → { TPH3006PD	600V	17A	0.15Ω	6nC
CMF20120	1200V	24A	0.098Ω	49nC

Current Research in Wide Bandgap Materials



} Research

Srabanti Chowdhury *et al* "Current status and scope of gallium nitride-based vertical transistors for high-power electronics application" 2013 *Semicond. Sci. Technol.* 28, 074014



MOSFET: conclusions

- A majority-carrier device: fast switching speed
- Typical switching frequencies: tens and hundreds of kHz & above MHz
- On-resistance increases rapidly with rated blocking voltage
- Easy to drive
- The device of choice for blocking voltages less than 500V
- 1000V devices are available, but are useful only at low power levels (100W)
- Part number is selected on the basis of on-resistance rather than current rating

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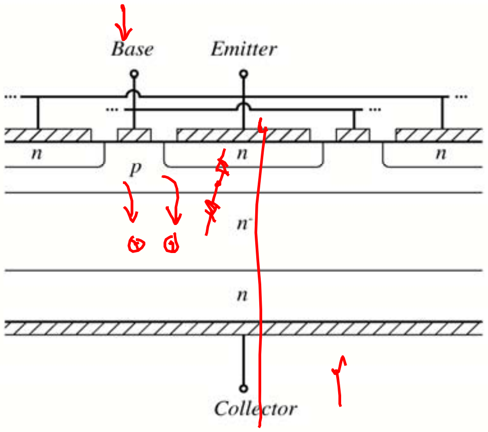
4.2.3. Bipolar Junction Transistor (BJT)



- Interdigitated base and emitter contacts
- Vertical current flow
- npn device is shown
- minority carrier device
- on-state: base-emitter and collector-base junctions are both forward-biased
- on-state: substantial minority charge in p and n^- regions, conductivity modulation

Conclusions: BJT

- BJT has been replaced by MOSFET in low-voltage (<500V) applications
- BJT is being replaced by IGBT in applications at voltages above 500V
- A minority-carrier device: compared with MOSFET, the BJT exhibits slower switching, but lower on-resistance at high voltages

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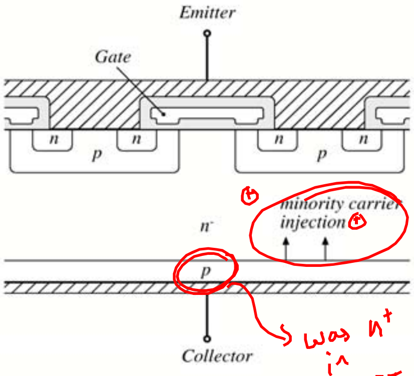
4.2.4. The Insulated Gate Bipolar Transistor (IGBT)

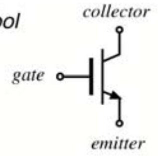


was n^+ in MOSFET

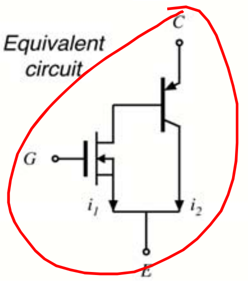
- A four-layer device
- Similar in construction to MOSFET, except extra *p* region
- On-state: minority carriers are injected into n^- region, leading to conductivity modulation
- compared with MOSFET: slower switching times, lower on-resistance, useful at higher voltages (up to 1700V)

The IGBT

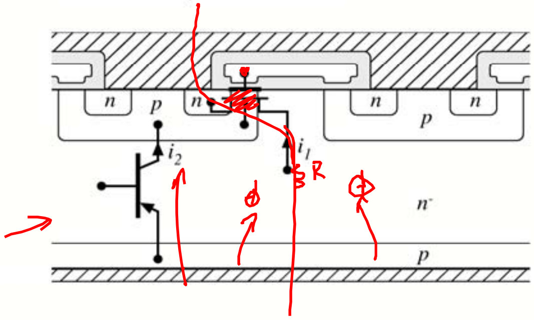
Symbol



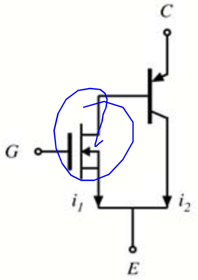
Equivalent circuit

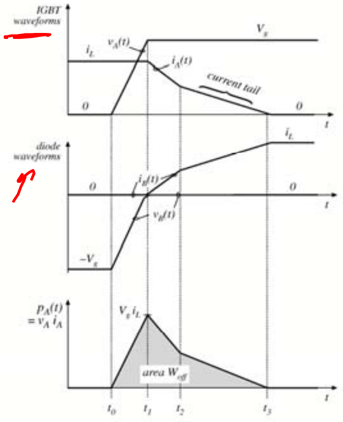


Location of equivalent devices

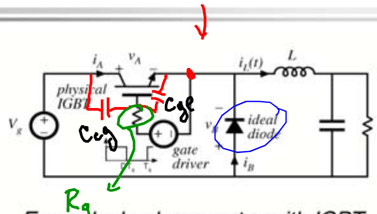


Current tailing in IGBTs

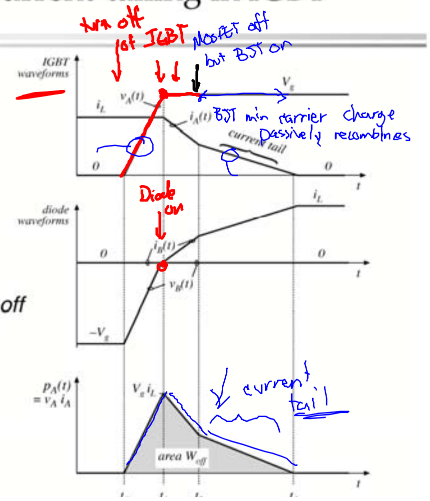




Switching loss due to current-tailing in IGBT



Example: buck converter with IGBT



transistor turn-off transition

$$P_{sw} = \frac{1}{T_s}\int_{\substack{\text{switching}\\ \text{transitions}}} p_A(t)\,dt = (W_{on} + W_{off})\,f_s$$

Characteristics of several commercial devices

Part number	Rated max voltage	Rated avg current	V_F (typical)	t_f (typical)
Single-chip devices				
HGTG32N60E2	600V	32A	2.4V	0.62μs
HGTG30N120D2	1200V	30A	3.2A	0.58μs
Multiple-chip power modules				
CM400HA-12E	600V	400A	2.7V	0.3μs
CM300HA-24E	1200V	300A	2.7V	0.3μs

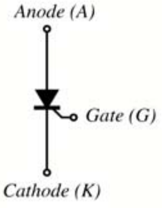
Conclusions: IGBT

- Becoming the device of choice in 500 to 1700V+ applications, at power levels of 1-1000kW
- Positive temperature coefficient at high current —easy to parallel and construct modules
- Forward voltage drop: diode in series with on-resistance. 2-4V typical
- Easy to drive —similar to MOSFET
- Slower than MOSFET, but faster than Darlington, GTO, SCR
- Typical switching frequencies: 3-30kHz
- IGBT technology is rapidly advancing:
    - 3300 V devices: HVIGBTs
    - 150 kHz switching frequencies in 600 V devices

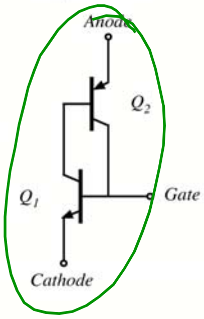
4.2.5. Thyristors (SCR, GTO, MCT)

The SCR

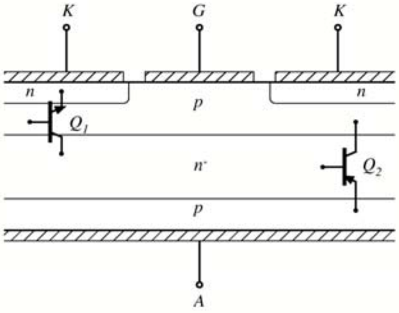
symbol



equiv circuit

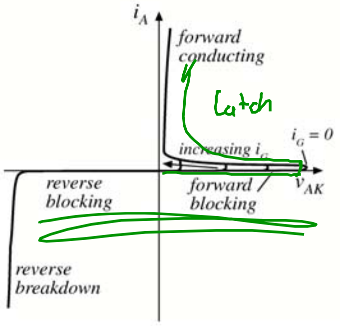


construction



The Silicon Controlled Rectifier (SCR)

- Positive feedback —a latching device
- A minority carrier device
- Double injection leads to very low on-resistance, hence low forward voltage drops attainable in very high voltage devices
- Simple construction, with large feature size
- Cannot be actively turned off
- A voltage-bidirectional two-quadrant switch
- 5000-6000V, 1000-2000A devices



Transistor Selection

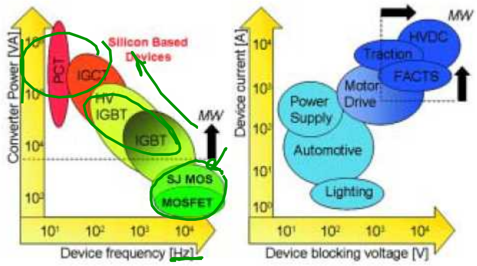
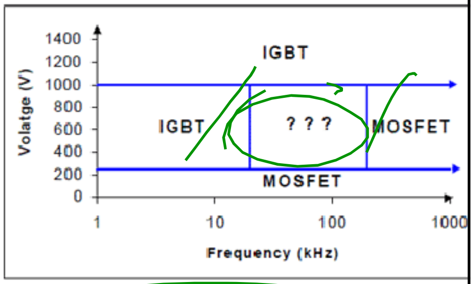


Figure 1: Power Semiconductor devices and applications



Transistor Selection

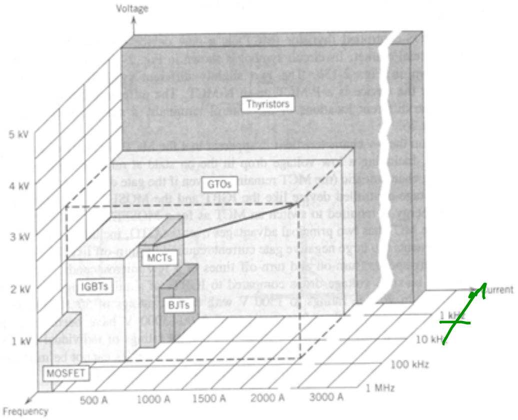
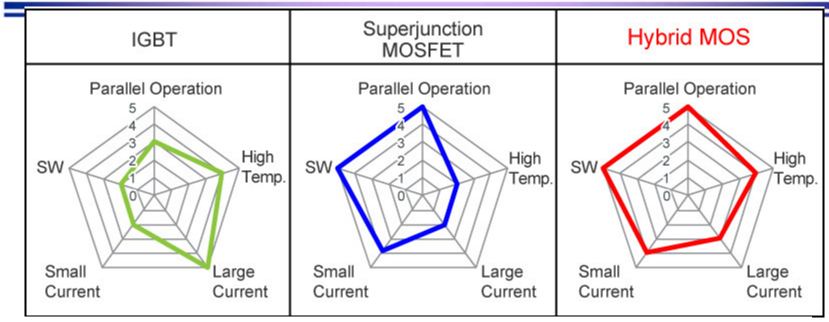
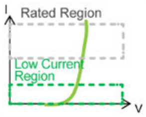
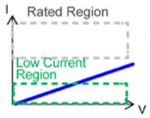
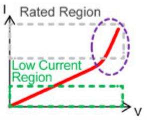


Figure 2-14 Summary of power semiconductor device capabilities. All devices except the MCT have a relatively mature technology, and only evolutionary improvements in the device capabilities are anticipated in the next few years. However, MCT technology is in a state of rapid expansion, and significant improvements in the device capabilities are possible, as indicated by the expansion arrow in the diagram.

Mohan

Transistor Hybridization



	IGBT	Superjunction MOSFET	Hybrid MOS
V-I Characteristics			
Low Current	×	○	○
High Current	○	×	△
High-Speed Switching	△	○	○

Significantly lower yearly power consumption

Excellent for steady-state operation

Ideal for instantaneous large currents

4.3. Switching loss

- Energy is lost during the semiconductor switching transitions, via several mechanisms:
  - ~~Transistor switching times~~ Outdated!
  - Diode stored charge Q_r
  - Energy stored in device capacitances and parasitic inductances
- Semiconductor devices are *charge controlled*
- Time required to insert or remove the controlling charge determines switching times

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Some other sources of this type of switching loss

Schottky diode (majority carrier)

- Essentially no stored charge $Q_r = 0$ but $C_j > 0$
- Significant reverse-biased junction capacitance

Transformer leakage inductance

- Effective inductances in series with windings
- A significant loss when windings are not tightly coupled

Interconnection and package inductances

- Diodes
- Transistors
- A significant loss in high current applications

4.3.4. Efficiency vs. switching frequency

Add up all of the energies lost during the switching transitions of one switching period:

$$W_{tot} = W_{on} + W_{off} + W_D + W_C + W_L + \ldots$$

Average switching power loss is

$$P_{sw} = W_{tot}\, f_{sw}$$

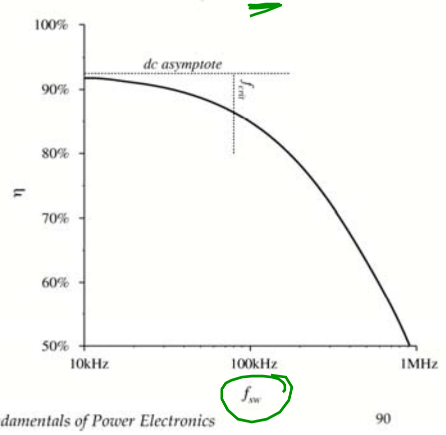
Total converter loss can be expressed as

$$P_{loss} = P_{cond} + P_{fixed} + W_{tot}\, f_{sw}$$

where P_{fixed} = fixed losses (independent of load and f_{sw})
P_{cond} = conduction losses

Efficiency vs. switching frequency

$$P_{loss} = P_{cond} + P_{fixed} + W_{tot}\, f_{sw}$$



Inclusion of Switching Loss in the Averaged Equivalent Circuit Model

The methods of Chapter 3 can be extended to include switching loss in the converter equivalent circuit model

- Include switching transitions in the converter waveforms
- Model effects of diode reverse recovery, etc.

To obtain tractable results, the waveforms during the switching transitions must usually be approximated

Things that can substantially change the results:

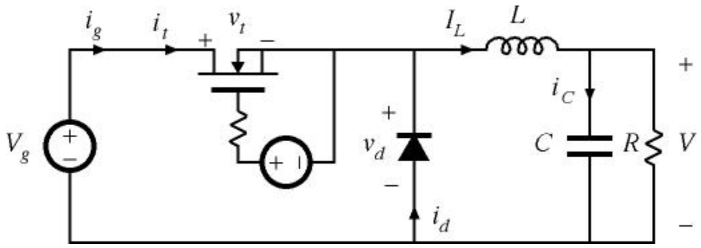
- Ringing caused by parasitic tank circuits
- Snubber circuits

The Modeling Approach

Extension of Chapter 3 Methods

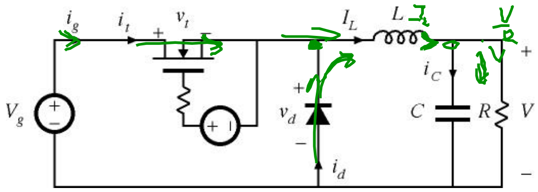
- Sketch the converter waveforms
  - Including the switching transitions (idealizing assumptions are made to lead to tractable results)
  - In particular, sketch inductor voltage, capacitor current, and input current waveforms
- The usual steady-state relationships:
  - $\langle v_L \rangle = 0, \langle i_C \rangle = 0, \langle i_g \rangle = I_g$
- Use the resulting equations to construct an equivalent circuit model, as usual

Buck Converter Example



- Ideal MOSFET, *p–n* diode with reverse recovery
- Neglect semiconductor device capacitances, MOSFET switching times, etc.
- Neglect conduction losses
- Neglect ripple in inductor current and capacitor voltage

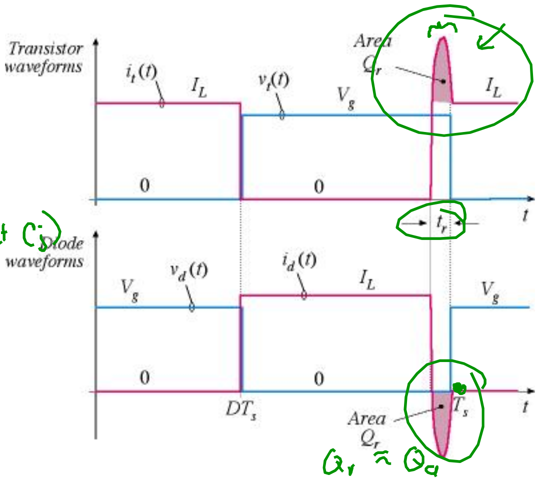
Assumed waveforms



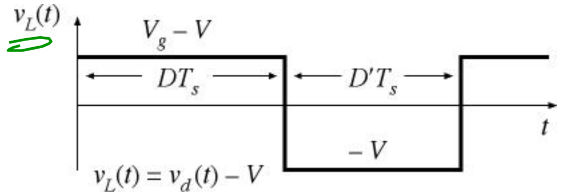
Diode recovered charge Q_r, reverse recovery time t_r

These waveforms assume that the diode voltage changes at the end of the reverse recovery transient

- a "snappy" diode (Neglect C_j)
- Voltage of soft-recovery diodes changes sooner
- Leads to a pessimistic estimate of induced switching loss



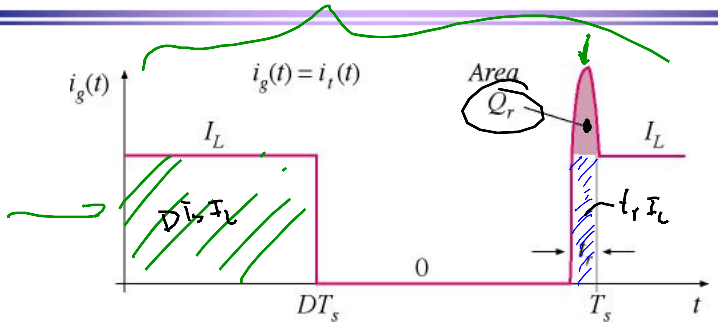
Inductor volt-second balance and capacitor charge balance



As usual: $\langle v_L \rangle = 0 = DV_g - V$

Also as usual: $\boxed{\langle i_C \rangle = 0 = I_L - V/R}$

Average input current



$$\langle i_g \rangle = I_g = (\text{area under curve})/T_s$$

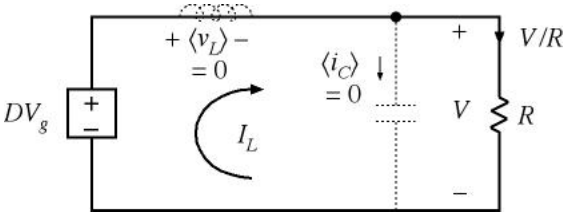
$$= (DT_sI_L + t_rI_L + Q_r)/T_s$$

$$\boxed{= DI_L + t_rI_L/T_s + Q_r/T_s} = \langle i_g \rangle$$

Construction of Equivalent Circuit Model

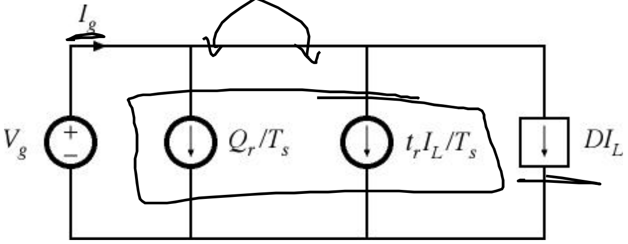
From inductor volt-second balance: $\langle v_L \rangle = 0 = DV_g - V$

From capacitor charge balance: $\langle i_C \rangle = 0 = I_L - V/R$

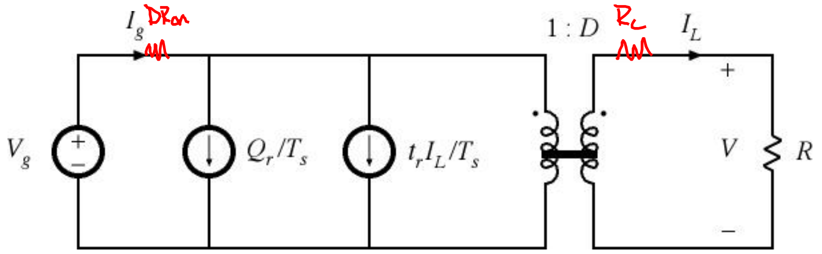


Input port of model

$$\langle i_g \rangle = I_g = DI_L + t_r I_L / T_s + Q_r / T_s$$



Combine for complete model



The two independent current sources consume power

$$V_g\,(t_r I_L/T_s + Q_r/T_s)$$

equal to the switching loss induced by diode reverse recovery

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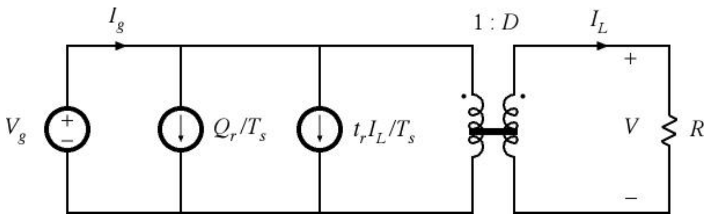
Solution of model

Output:

$V = DV_g$



Efficiency: $\eta = P_{out} / P_{in}$

$P_{out} = VI_L$ $\quad P_{in} = V_g\,(DI_L + t_r I_L/T_s + Q_r/T_s)$

Combine and simplify:

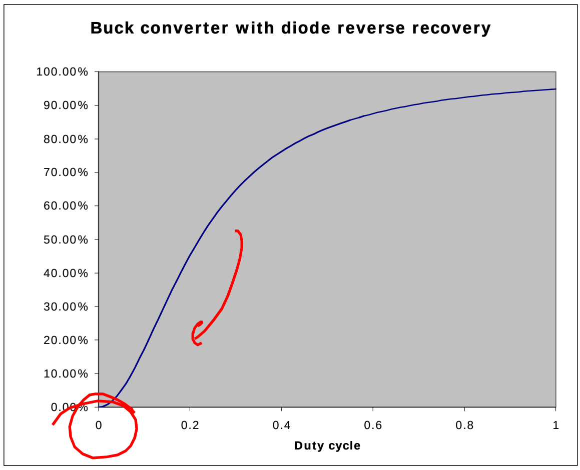
$\eta = 1 / [1 + f_s\,(t_r/D + Q_r R/D^2 V_g)]$ $= \dfrac{1}{1 + f_s(\;)}$

Predicted Efficiency vs Duty Cycle

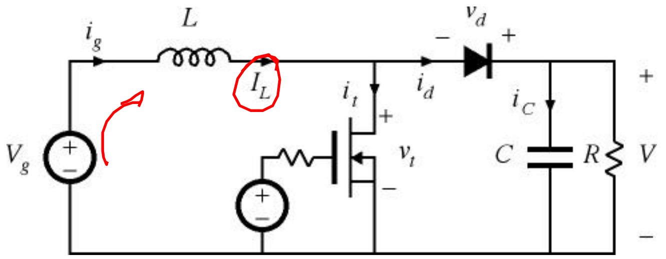
- Switching frequency 100 kHz
- Input voltage 24 V
- Load resistance 15 Ω
- Recovered charge 0.75 µCoul
- Reverse recovery time 75 nsec

- (no attempt is made here to model how the reverse recovery process varies with inductor current)

- Substantial degradation of efficiency
- Poor efficiency at low duty cycle

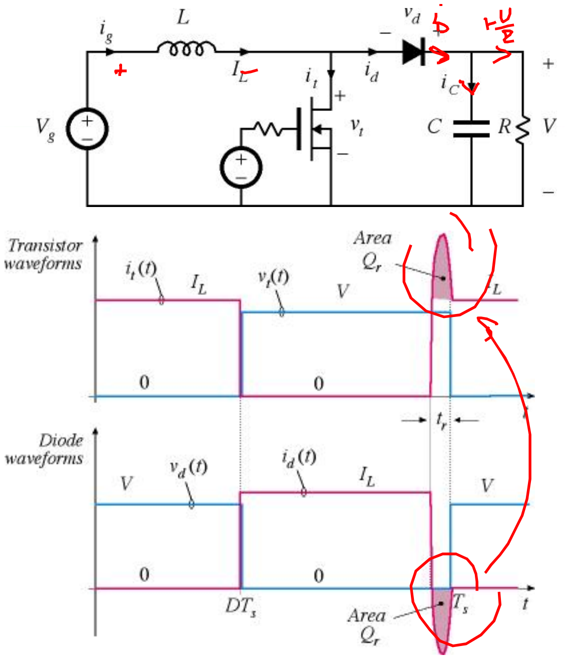


Boost Converter Example



- Model same effects as in previous buck converter example:
- Ideal MOSFET, *p–n* diode with reverse recovery
- Neglect semiconductor device capacitances, MOSFET switching times, etc.
- Neglect conduction losses
- Neglect ripple in inductor current and capacitor voltage

Boost converter



Transistor and diode waveforms have same shapes as in buck example, but depend on different quantities

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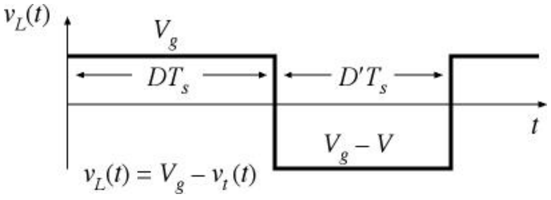
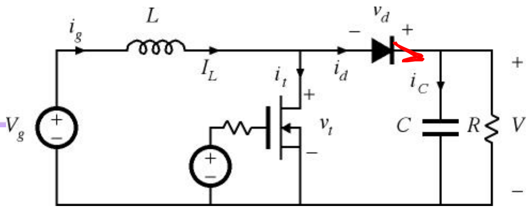
Inductor volt-second balance and average input current

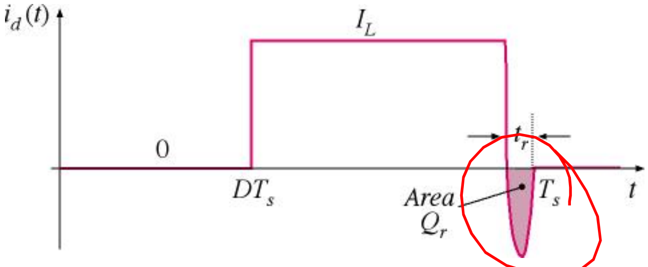


As usual: $\langle v_L \rangle = 0 = V_g - D'V$

Also as usual: $\langle i_g \rangle = I_L$

Capacitor charge balance





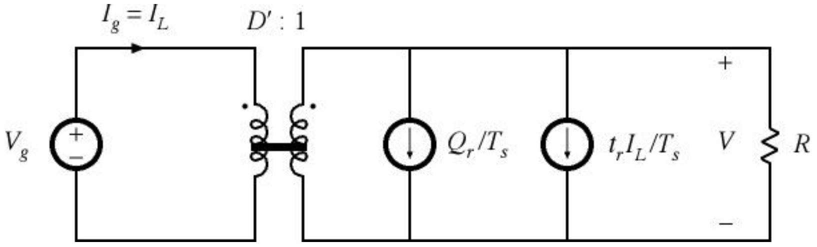
$$\langle\, i_C \,\rangle = \langle\, i_d \,\rangle - V/R = 0$$

$$= -\,V/R + I_L(D'T_s - t_r)/T_s - Q_r\,/T_s$$

Collect terms: $V/R = \underbrace{I_L(D'T_s - t_r)/T_s - Q_r\,/T_s}$

Construct model

The result is:



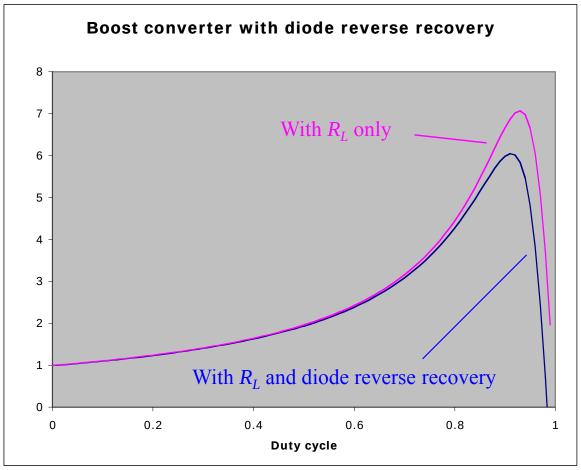
The two independent current sources consume power

$$V\,(t_rI_L\,/T_s + Q_r\,/T_s)$$

equal to the switching loss induced by diode reverse recovery

Predicted V/V_g vs duty cycle

- Switching frequency 100 kHz
- Input voltage 24 V
- Load resistance 60 Ω
- Recovered charge 5 μCoul
- Reverse recovery time 100 nsec
- Inductor resistance $R_L = 0.3\ \Omega$
- (inductor resistance also inserted into averaged model here)



Summary

- The averaged modeling approach can be extended to include effects of switching loss
- Transistor and diode waveforms are constructed, including the switching transitions. The effects of the switching transitions on the inductor, capacitor, and input current waveforms can then be determined
- Inductor volt-second balance and capacitor charge balance are applied
- Converter input current is averaged
- Equivalent circuit corresponding to the the averaged equations is constructed