



University
of Glasgow

Power Electronics

Lecture 5

Power Switches

For Switching Power Conversion

Please read
pages 16-31 in Chapter 2
Pages 546-661 Chapter 21-26
of the textbook

Typical Controllable Power Switches.

There are several different electronic devices used for switching appreciable amounts of electrical power. Those to be examined in the next two lectures are:

1. Thyristors (Silicon Controlled Rectifiers; SCRs) 晶闸管 (half-controlled)
2. GTOs (Gate turn-off Thyristors) 可关断晶闸管 (fully controlled)
3. TRIACs (triode for alternating current) 双向晶闸管 (half-controlled)
4. MOSFETs (Metal Oxide Semiconductor Field Effect Transistors) (fully controlled)
5. GTR (Giant Transistor) 大功率晶体管 (fully controlled)
6. IGBT (Insulated Gate Bipolar Transistors) 绝缘栅极双极性晶体管 (fully controlled)



Practical Considerations for Switch Selection:

I. How much control do we have (or need)?

→ Controllability

I. How much voltage will it drop when on?

→ Conduction Resistance

I. How much voltage can it block 阻断 when off?

→ Voltage Rating

I. How much current can it carry 承受?

→ Current Rating

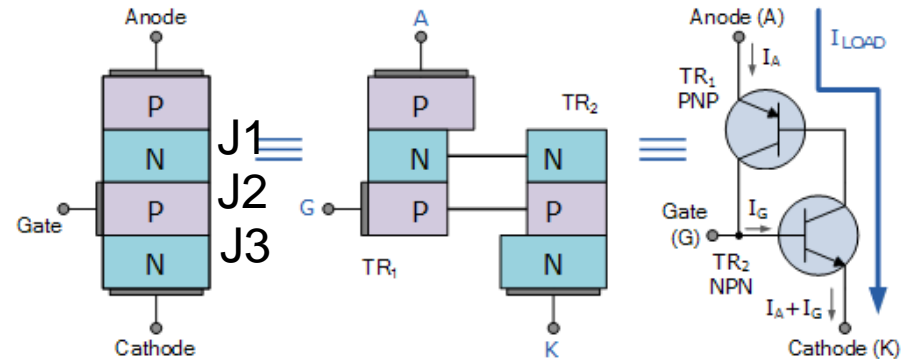
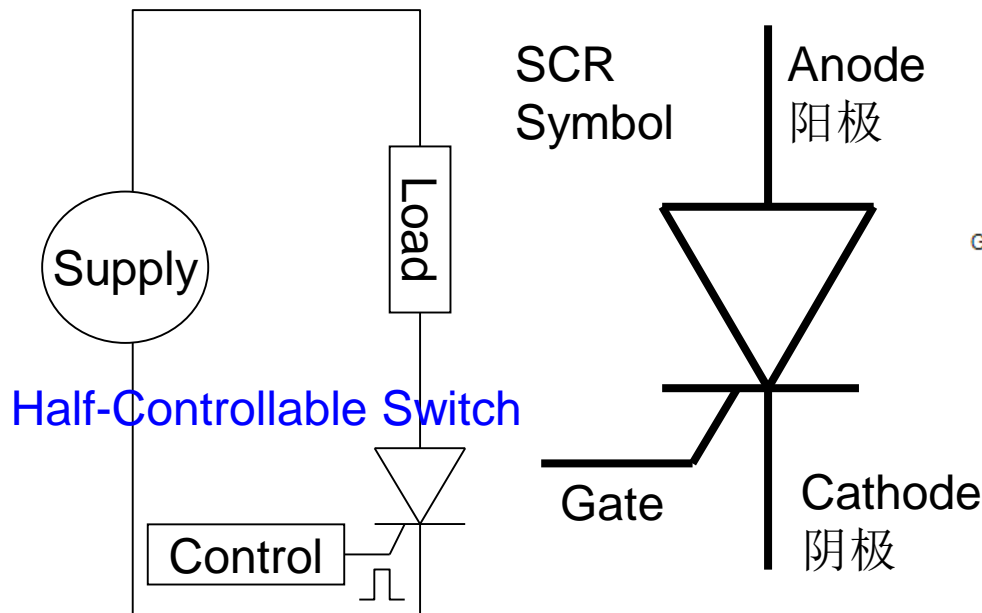
I. How long does it take to turn on and off?

→ Switching Frequency



The Thyristor (SCR – Silicon-Controlled-Rectifier) 晶闸管

If the anode voltage is higher than the cathode voltage, a brief pulse on the gate will turn on the device. Once the device is conducting, it stays on (“Latches” 锁) until the current through it falls to zero. Once on, the gate loses all control.



$$I_A = \frac{\alpha_2 I_G + I_{CO1} + I_{CO2}}{1 - (\alpha_1 + \alpha_2)}$$

The gate needs a few volts and a few milliamps to switch the device on.

The device can pass thousands of amps and block thousands of volts.

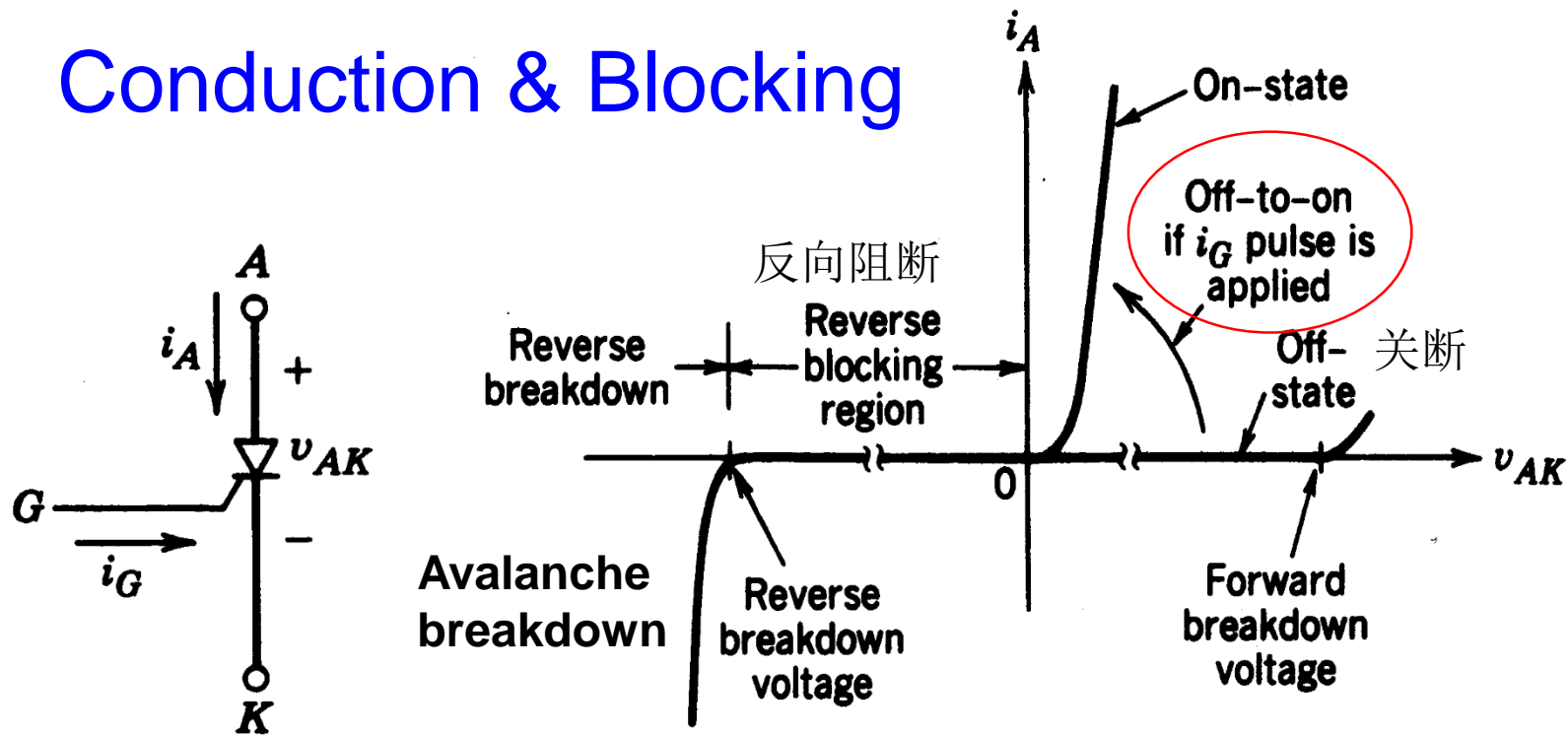
When on, the forward voltage drop is low (Approximate 1V).

Turn-on is slow, taking a few 10's to a few 100's of microseconds, depending on the size of the device.

- ◆ 晶体管的特性是：在低发射极电流下 α 是很小的，而当发射极电流建立起来之后， α 迅速增大。
- ◆ 在晶体管阻断状态下， $I_G=0$ ，而 $\alpha_1+\alpha_2$ 是很小的。由上式可看出，此时流过晶闸管的漏电流只是稍大于两个晶体管漏电流之和。
- ◆ 如果注入触发电流使各个晶体管的发射极电流增大以致 $\alpha_1+\alpha_2$ 趋近于1的话，流过晶闸管的电流 I_A （阳极电流）将趋近于无穷大，从而实现器件饱和导通。
- ◆ 由于外电路负载的限制， I_A 实际上会维持有限值。

They are used in very high power applications!!!

Conduction & Blocking



Half-Controllable Switch 半控开关

The thyristor can be turned on by applying a pulse of positive gate current i_G for a short duration when the device is in its forward-blocking 前向阻断 state.

Once the device begins to conduct, it is latched on and i_G can be removed. It cannot be turned off by gate. It will turn off until anode current i_A goes to zero.



Trigger On

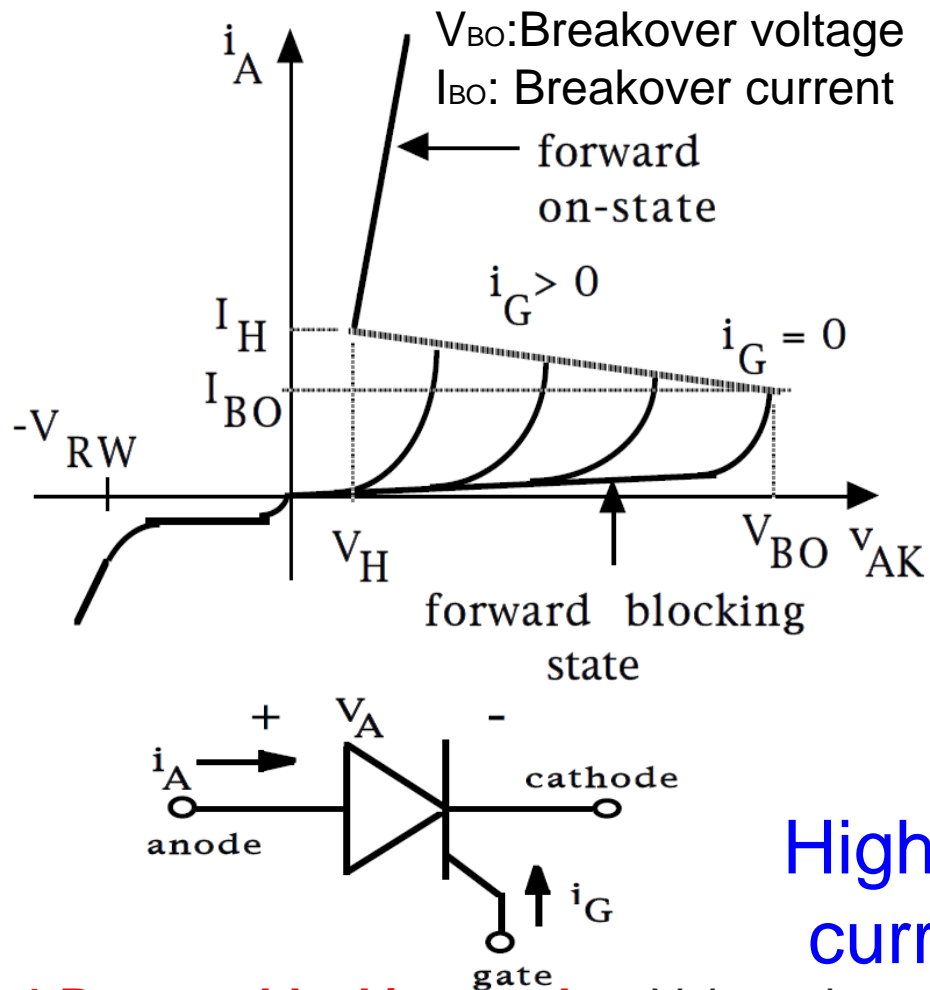
Current to **several kiloamps** for $V_{(on)}$ of about **1 volts**.

Blocking voltages to **5-8 kilovolts**.

Maximum junction temperature = **125 °C**

Higher i_G leads to shorter turn-on time

High efficient for high voltage high current applications (e.g. HVDC)

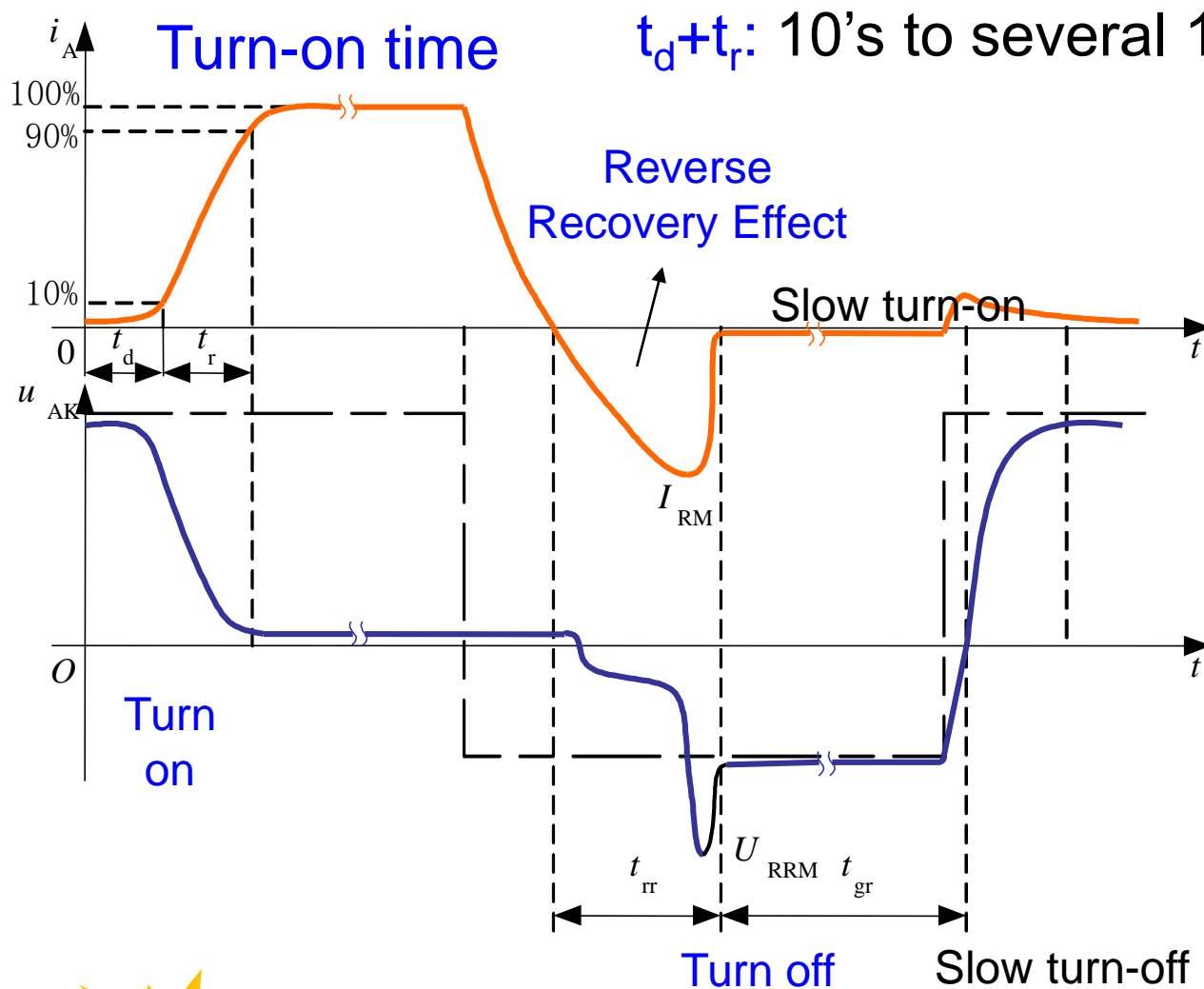


1.Reverse blocking mode – Voltage is applied in the direction that would be blocked by a diode

2.Forward blocking mode – Voltage is applied in the direction that would cause a diode to conduct, but the thyristor has not been triggered into conduction

3.Forward conducting mode – The thyristor has been triggered into conduction and will remain conducting until the forward current drops below a threshold value known as the "holding current" I_H which corresponds the holding voltage V_H

Turn-on/off of Thyristor (SCR)



After the reverse recovery process is completed, because the carrier recombination process is relatively slow, it will take some time for the thyristor to recover its ability to block forward voltage. This is called the forward blocking recovery time.

If a forward voltage is applied to the thyristor again during the forward blocking recovery time, the thyristor will re-conduct, instead of conducting under gate current control.

t_{gr} : 正向阻断恢复时间

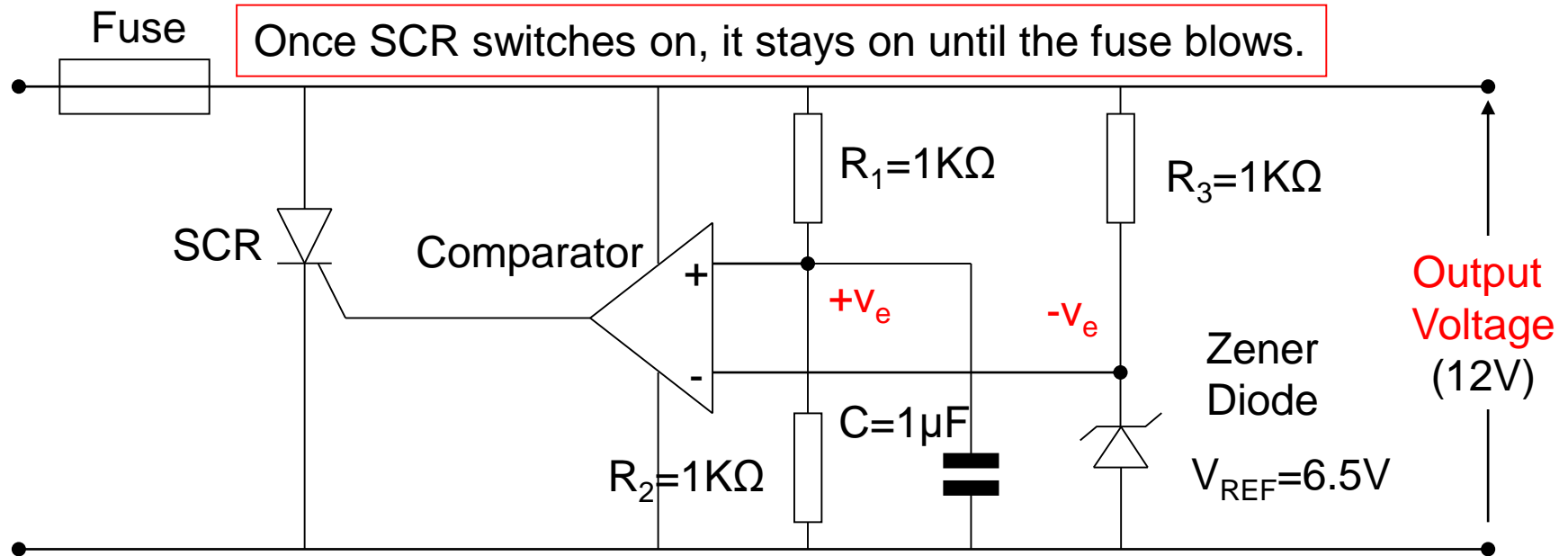
t_{rr} : 反向阻断恢复时间



Turn-off time $t_{rr} + t_{gr}$: several hundred μs

Example application of SCRs in DC circuits

The diagram below shows a power supply with an SCR used to prevent the output from exceeding a pre-defined value.



This is called an *over-voltage crowbar circuit*.

Comparator truth table

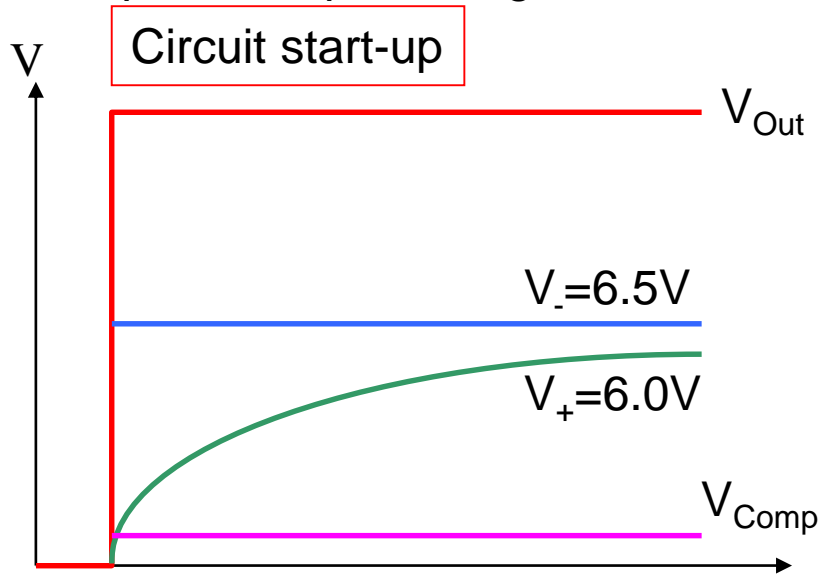
$+ve > -ve$	Output \uparrow
$+ve < -ve$	Output \downarrow

If the comparator + input is more positive than the – input, the output (V_{Comp}) is “on”.

If the comparator + input is more negative than the – input, the output is “off”.



Comparator input voltages.

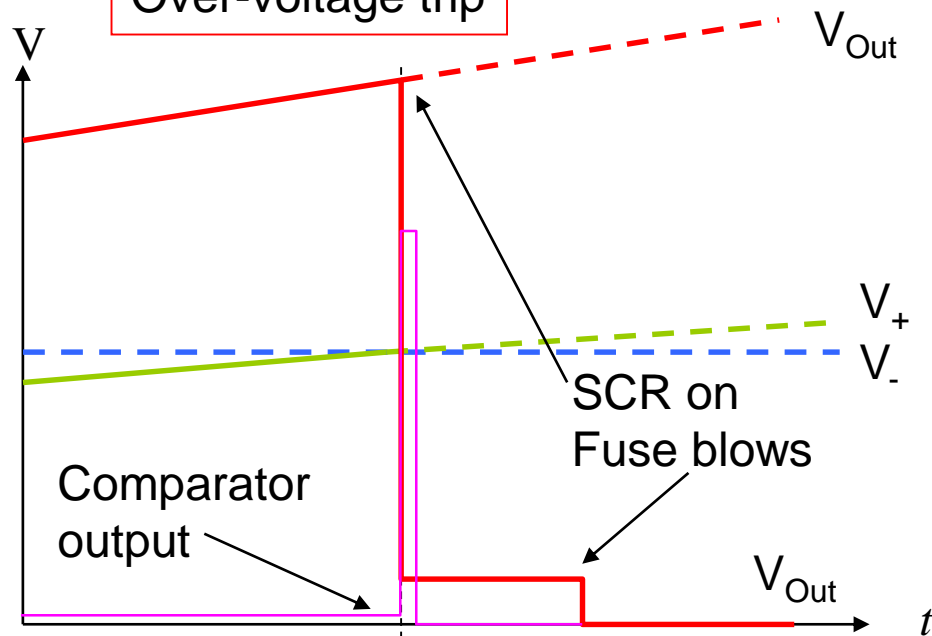


The capacitor holds the voltage on the +ve comparator input down whilst the supply voltages stabilise, preventing false triggering 触发 of the SCR. This also provides noise immunity when operating. $T=RC=1$ second.

R_1 & R_2 form a **potential divider** 分压电路 holding the +ve input at half the output voltage (6V).

R_3 quickly biases the **zener diode** 齐纳二极管 into the conduction region so that the -ve input is held at 6.5V.

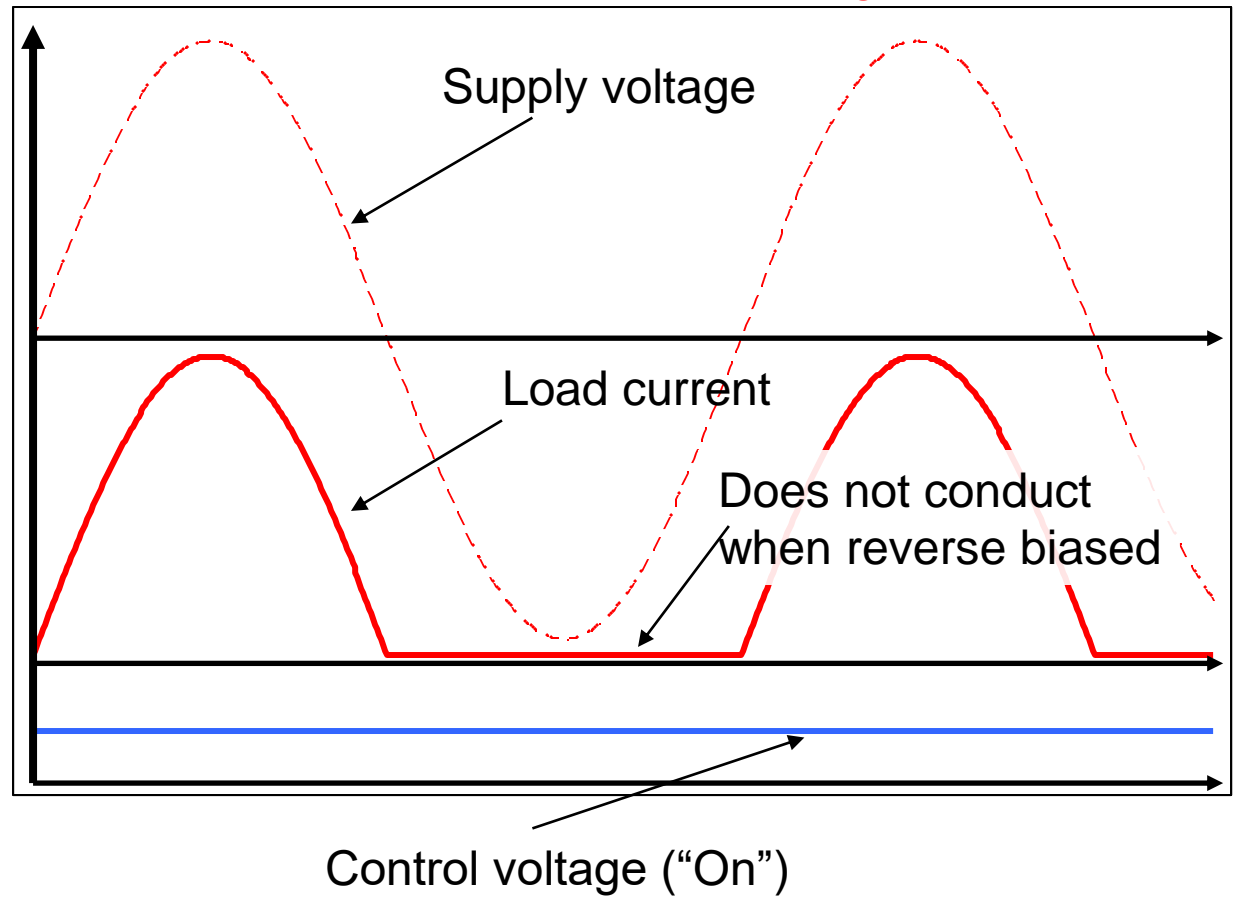
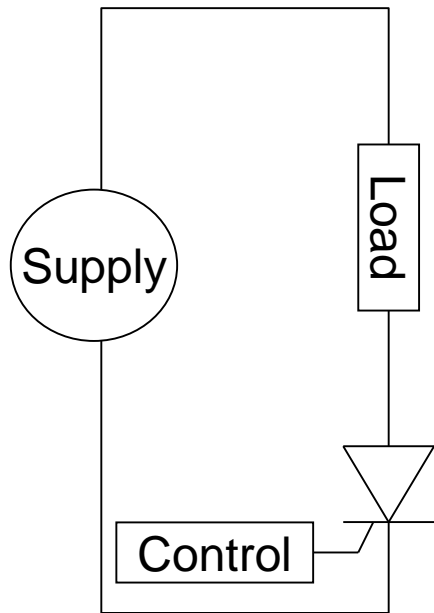
Over-voltage trip



If the output voltage starts to rise, the potential divider also starts to rise but the zener voltage is fixed. When V_+ is greater than V_- the comparator output switches on, fires the SCR which then latches on. This presents a short circuit across the supply, protecting the load. The fuse then **blows** 熔断 to protect the transformer etc. After the fuse blows the SCR resets itself.

Thyristor Switching

The previous showed how an SCR can be used as switch in a DC circuit to protect the equipment connected to the output of a power supply. This took advantage of the fact that once an SCR is on, it stays on until the current through it falls to zero. We would also like to be able to use SCRs in AC circuits because of their **huge power control capacity**.

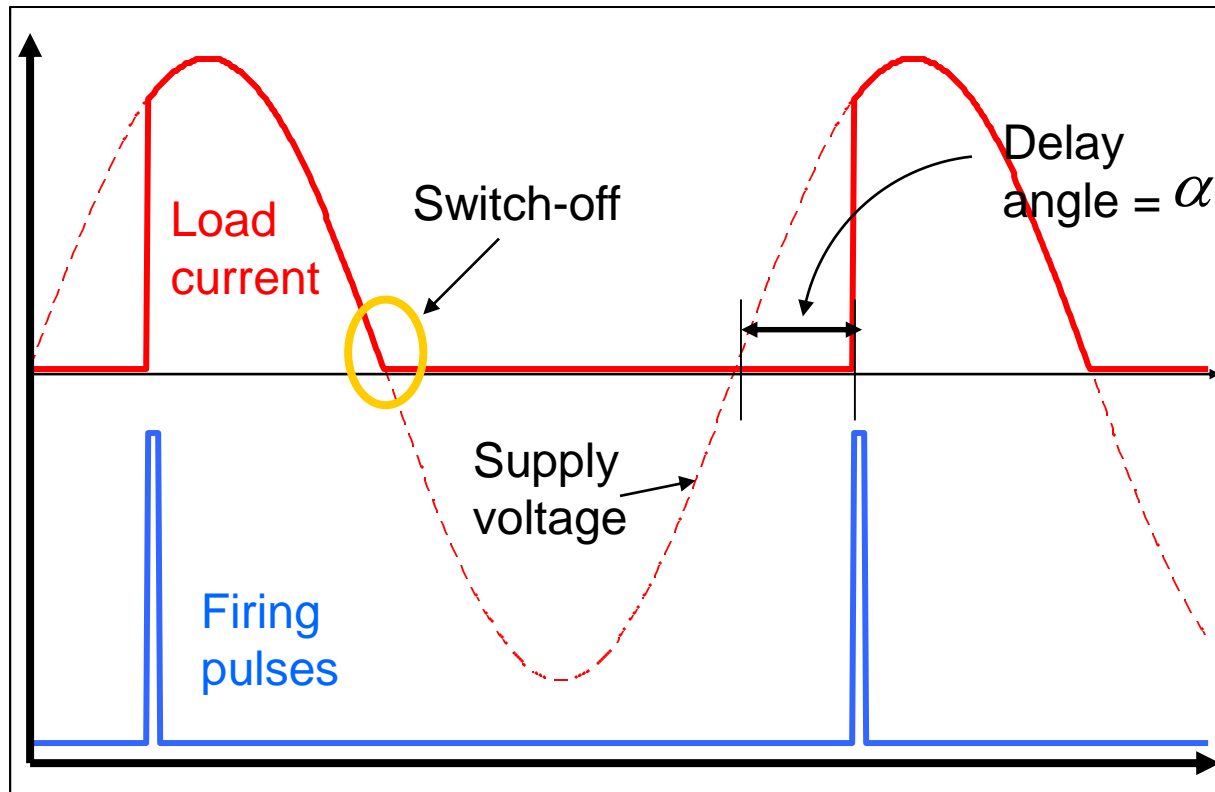


Thyristor behaving as a half-wave rectifier. The SCR turns off (“commutates” 换流) automatically at the end of each positive half cycle.



Phase control of SCRs

If we delay the firing pulse 触发脉冲 for the SCR relative to the zero-crossing points, we can use the SCR as a means 手段 of regulating the power applied to the load.



The period of time before the current flows in the SCR is called the delay angle, α



Calculation of average and RMS voltages for half-wave SCR phase angle control

$$\begin{aligned} V_{AVE} &= \frac{1}{2\pi} \left[\int_0^{\alpha} 0 \cdot d\omega t + \int_{\alpha}^{\pi} \hat{V} \sin(\omega t) d\omega t \right] \\ &= \frac{\hat{V}}{2\pi} \left[-\cos(\omega t) \right]_{\alpha}^{\pi} \\ &= \frac{\hat{V}}{2\pi} \left[-\cos(\pi) + \cos(\alpha) \right] \\ &= \frac{\hat{V}}{2\pi} \left[1 + \cos(\alpha) \right] \end{aligned} \quad \begin{aligned} V_{RMS}^2 &= \frac{1}{2\pi} \left[\int_0^{\alpha} 0 \cdot d\omega t + \int_{\alpha}^{\pi} \left(\hat{V} \sin(\omega t) \right)^2 d\omega t \right] \\ &= \frac{\hat{V}^2}{2\pi} \int_{\alpha}^{\pi} \left[\frac{1}{2} (1 - \cos(2\omega t)) \right] d\omega t \\ &= \frac{\hat{V}^2}{4\pi} \left[\omega t - \frac{1}{2} \sin(2\omega t) \right]_{\alpha}^{\pi} \\ &= \frac{\hat{V}^2}{4\pi} \left[\pi - \alpha + \frac{1}{2} \sin(2\alpha) \right] \end{aligned}$$

Examples

For $V_{MAX} = 100V$

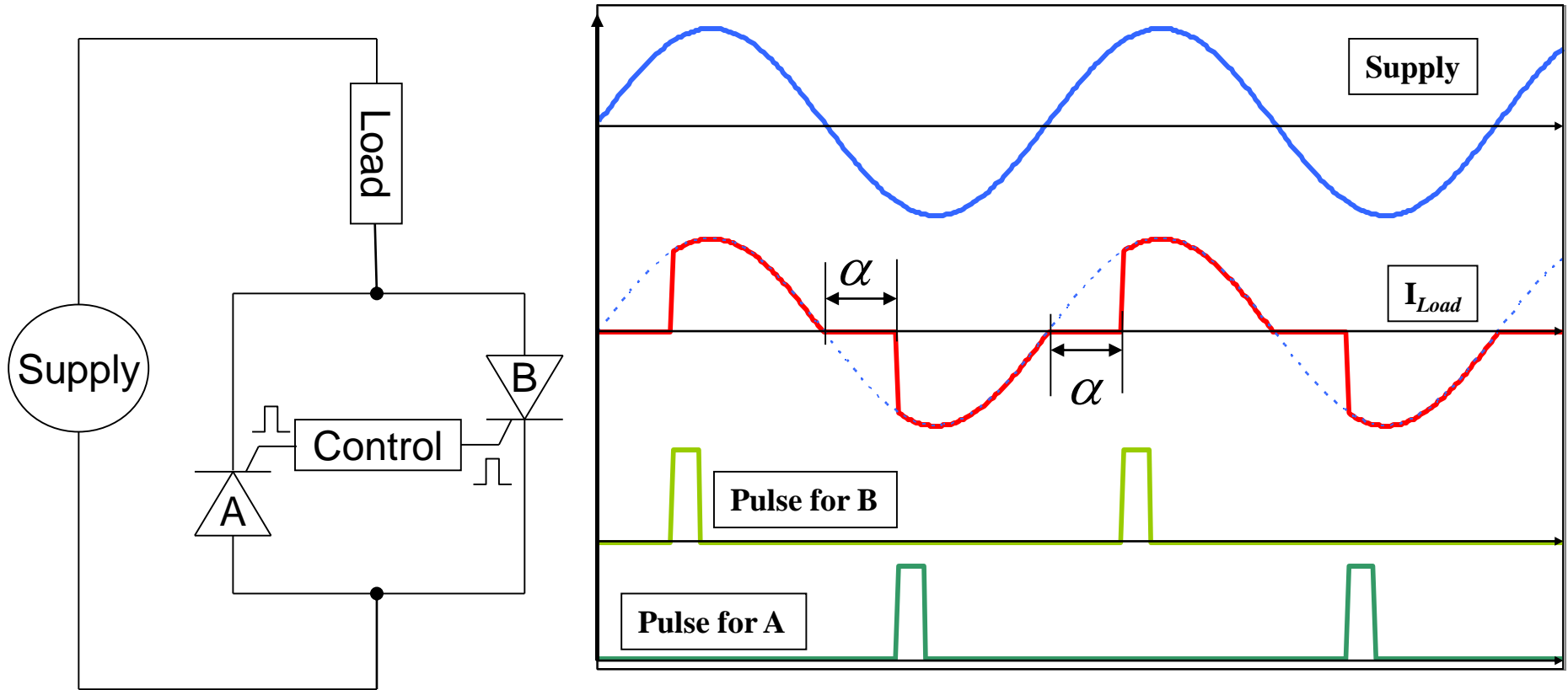
$$\alpha = 45^\circ: V_{AVE} = 27.17V, V_{RMS} = 47.67V$$

$$\alpha = 90^\circ: V_{AVE} = 15.92V, V_{RMS} = 35.36V$$

$$\alpha = 135^\circ: V_{AVE} = 4.66V, V_{RMS} = 15.07V$$

Back to back SCRs for full wave control

If we want to be able to control both halves of the mains waveform with SCRs we need to use a pair of SCRs.



By varying α we can change the firing point and therefore vary the power in the load.

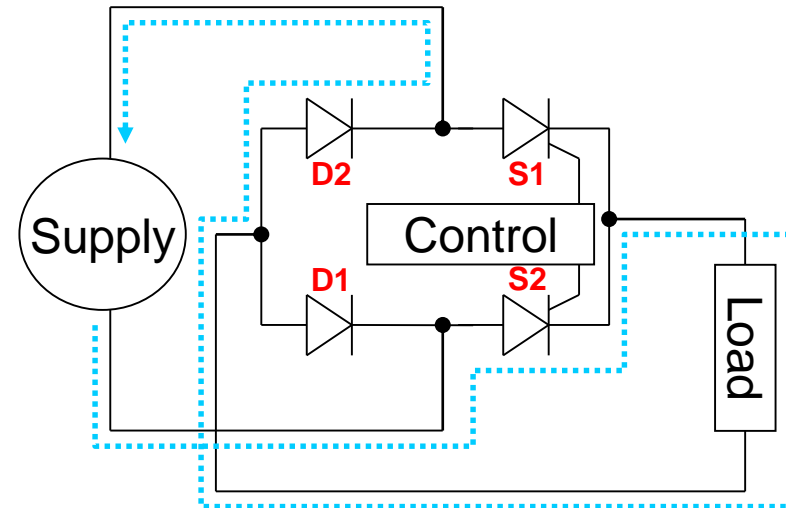
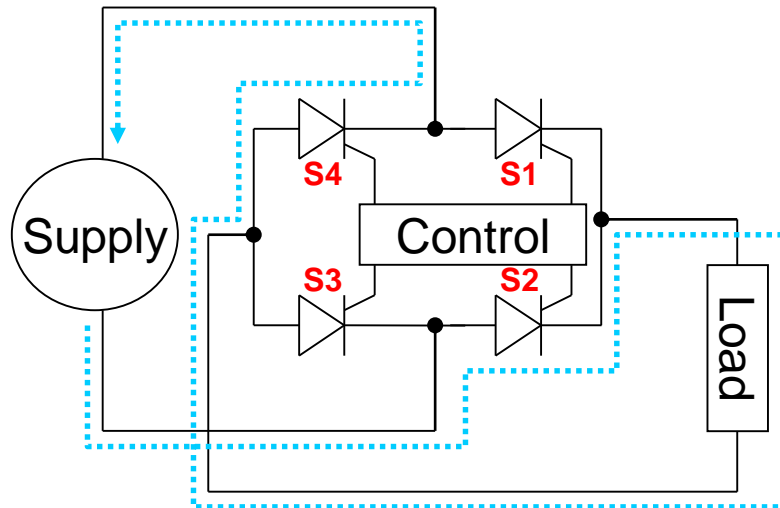
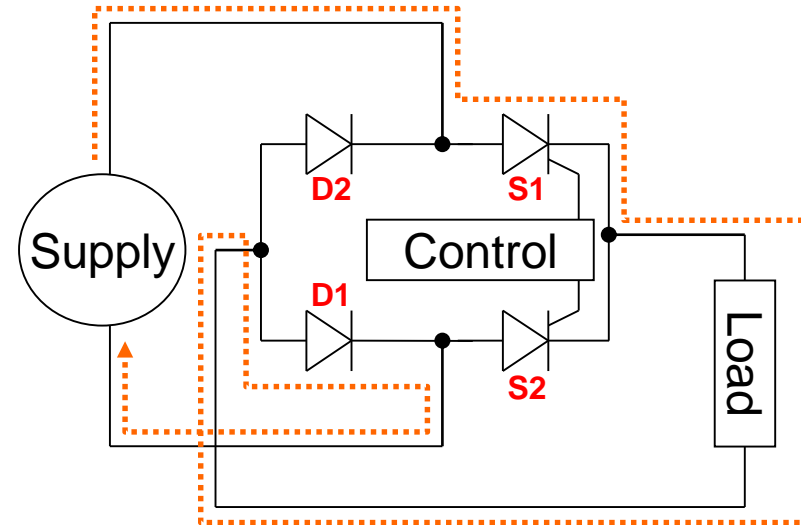
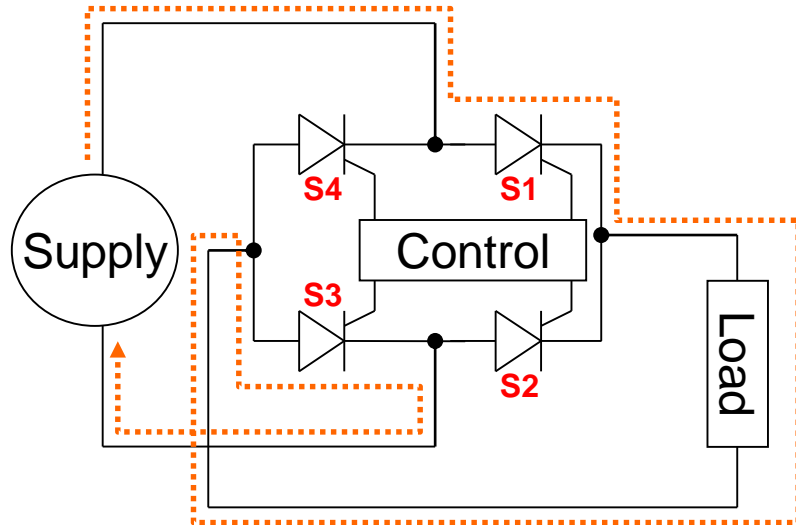
Calculation of average and RMS voltages for full-wave SCR phase angle control

$$\begin{aligned} V_{AVE} &= \frac{1}{\pi} \left[\int_0^{\alpha} 0 \cdot d\omega t + \int_{\alpha}^{\pi} \hat{V} \sin(\omega t) d\omega t \right] \\ &= \frac{\hat{V}}{\pi} \left[-\cos(\omega t) \right]_{\alpha}^{\pi} \\ &= \frac{\hat{V}}{\pi} \left[-\cos(\pi) + \cos(\alpha) \right] \\ &= \frac{\hat{V}}{\pi} \left[1 + \cos(\alpha) \right] ; 0 \leq \alpha \leq \pi \end{aligned}$$

$$\begin{aligned} V_{RMS}^2 &= \frac{1}{\pi} \left[\int_0^{\alpha} 0 \cdot d\omega t + \int_{\alpha}^{\pi} \left(\hat{V} \sin(\omega t) \right)^2 d\omega t \right] \\ &= \frac{\hat{V}^2}{\pi} \int_{\alpha}^{\pi} \left[\frac{1}{2} (1 - \cos(2\omega t)) \right] d\omega t \\ &= \frac{\hat{V}^2}{2\pi} \left[\omega t - \frac{1}{2} \sin(2\omega t) \right]_{\alpha}^{\pi} \\ &= \frac{\hat{V}^2}{2\pi} \left[\pi - \alpha + \frac{1}{2} \sin(2\alpha) \right] \end{aligned}$$

Thyristor Bridges

We can replace the diodes in a bridge rectifier with a pair of (or four) SCRs:



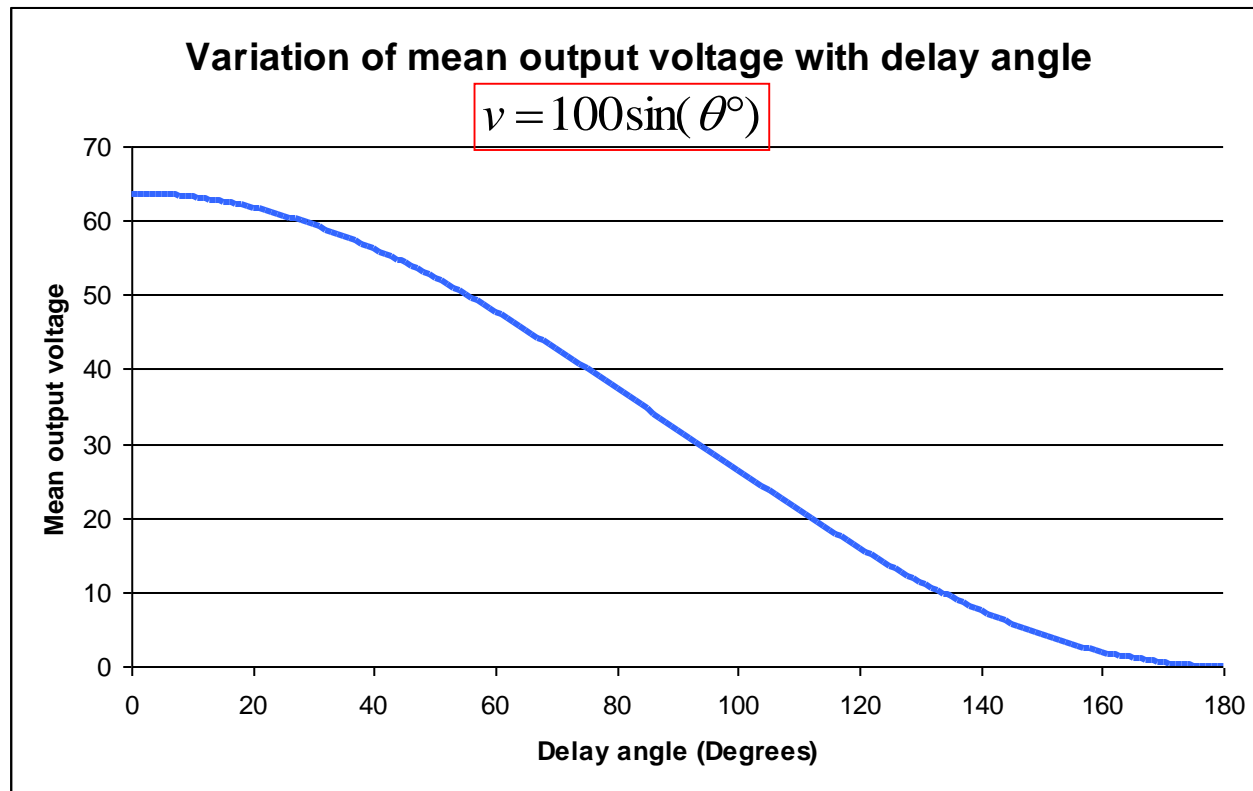
Positive & negative half-cycle current flows through SCRs and diodes.



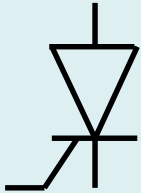
When we introduce phase control into the bridge, the average voltage is given by the integral of the waveform from the delay angle to the end of the half-cycle.

$$V_{AVE} = \frac{\hat{V}}{\pi} [1 + \cos(\alpha)] ; 0 \leq \alpha \leq \pi$$

The thyristor bridge allows us to vary the mean DC output voltage directly. The following chart of delay angle vs mean DC voltage is for a resistive load only.



Device selection chart

Device	Symbol	Control	Power Rating	Switch speed
SCR		Partial	Very high	Slow
TRIAC & SSR				
MOSFET				
IGBT				
GTO				

Example

If $V_{\text{Supply}} = 415V_{\text{RMS}}$ and the load is a resistor of value 25Ω , what power is dissipated in the load for delay angles of 30° , 90° and 150° ? What is the maximum load power when the SCRs are permanently switched on?

Maximum power = $V^2/R = 415^2 / 25 = 6.889\text{KW}$.

$$\text{For } \alpha = 30^\circ, V_{\text{RMS}}^2 = \frac{(415\sqrt{2})^2}{2\pi} \left[\pi - \frac{\pi}{6} + \frac{1}{2} \sin(60^\circ) \right] = 167259 .$$

$$P = \frac{V_{\text{RMS}}^2}{R} = 6690 \text{ W}$$

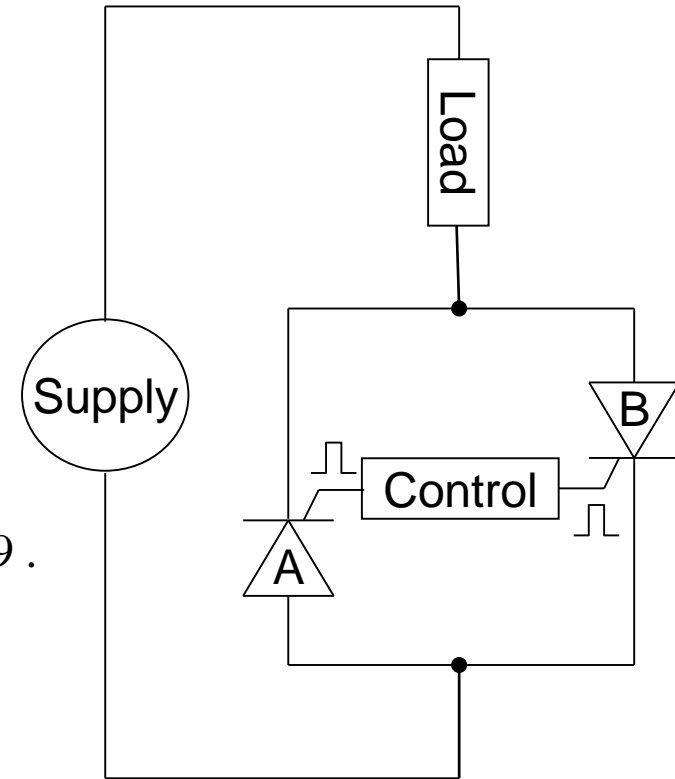
$$\text{For } \alpha = 90^\circ, V_{\text{RMS}}^2 = \frac{(415\sqrt{2})^2}{2\pi} \left[\pi - \frac{\pi}{2} + \frac{1}{2} \sin(180^\circ) \right] = 86112 .5$$

$$P = \frac{V_{\text{RMS}}^2}{R} = 3444 .5\text{W}$$

← This is exactly half the full power, which makes sense since the SCRs are on for exactly half the time.

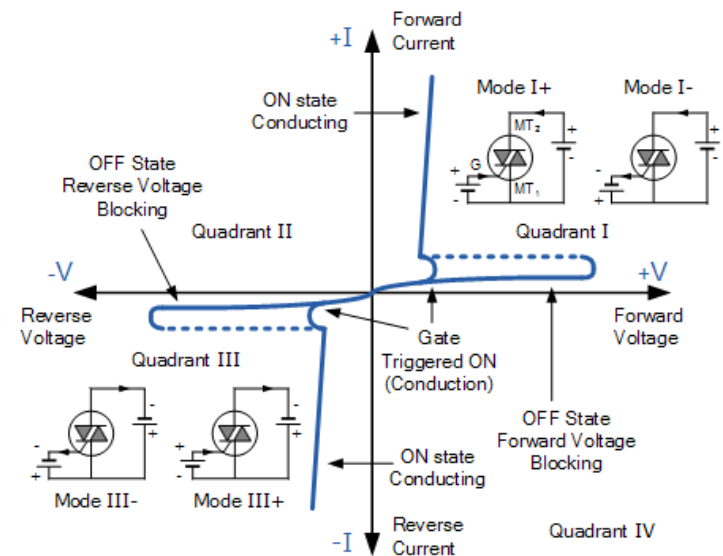
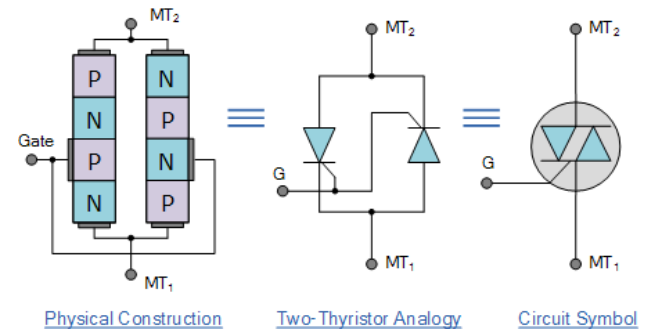
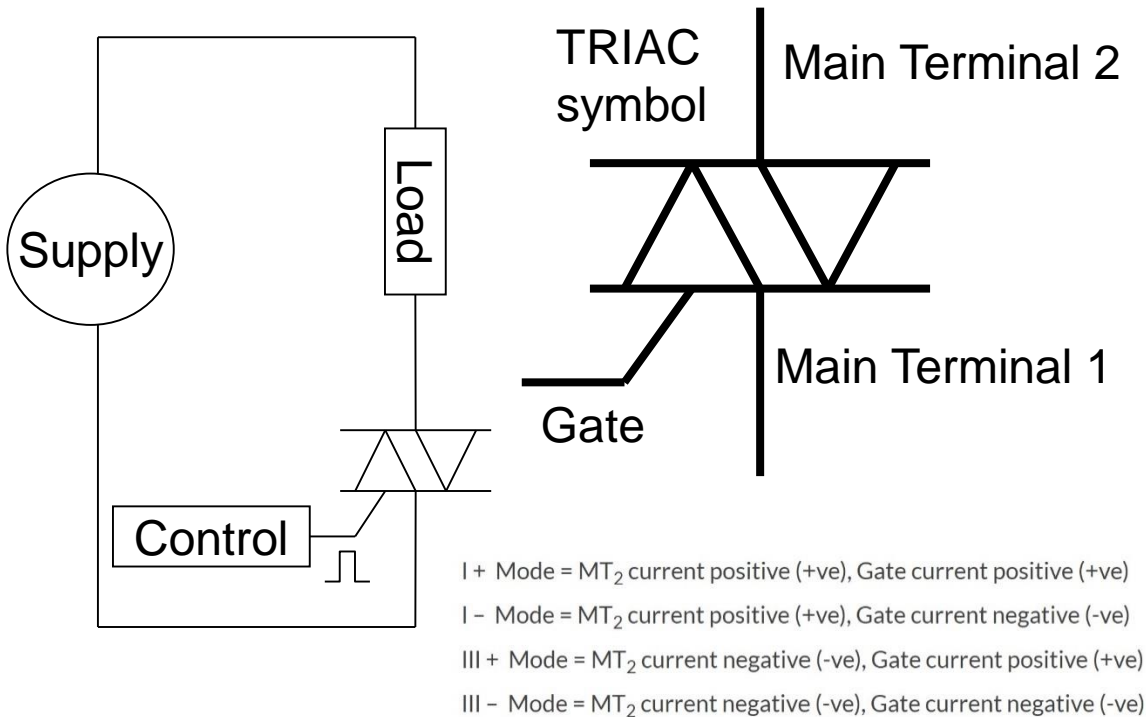
$$\text{For } \alpha = 150^\circ, V_{\text{RMS}}^2 = \frac{(415\sqrt{2})^2}{2\pi} \left[\pi - \frac{5\pi}{6} + \frac{1}{2} \sin(300^\circ) \right] = 4966$$

$$P = \frac{V_{\text{RMS}}^2}{R} = 198.6\text{W}$$



The TRIAC(triode for alternating current) 双向晶闸管

From a functional point of view the TRIAC is basically a pair of **back to back** 背靠背 SCRs sharing a common gate. As with the SCR, the TRIAC can be switched on by a short pulse applied to the gate. Once on, the device is latched until the current flowing through it falls to zero – usually the end of the AC half cycle.

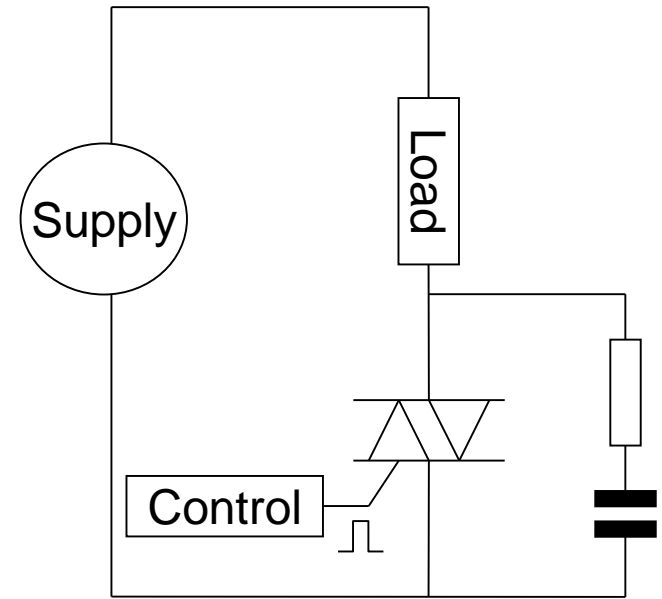
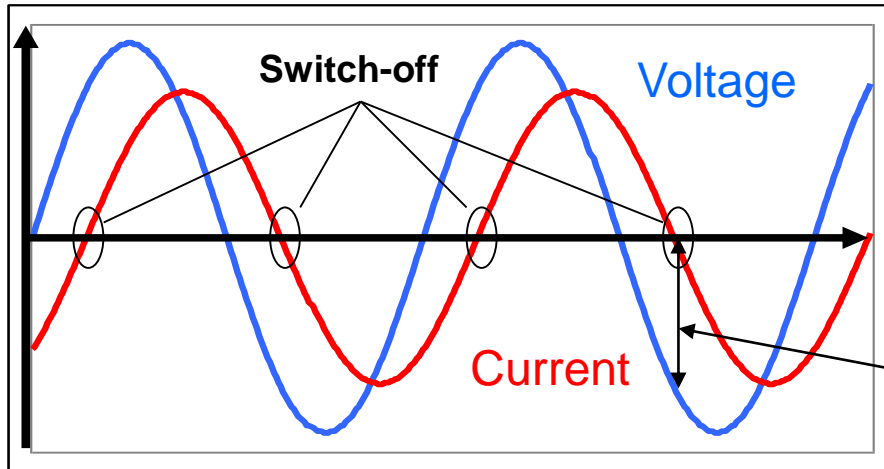


<https://www.electronics-tutorials.ws/power/triac.html>

TRIACs are bidirectional devices – they can pass current in either direction. They do not have such high power ratings as the SCR. However, they are very simple to use and find application in things like **dimmer** 调光 switches, heater controls and the like.

The TRIAC can be turned on by **either a positive or negative gate pulse**, irrespective of the polarity of the main terminal voltages. Turn-on times are of the order of a few microseconds. The triggering modes are:

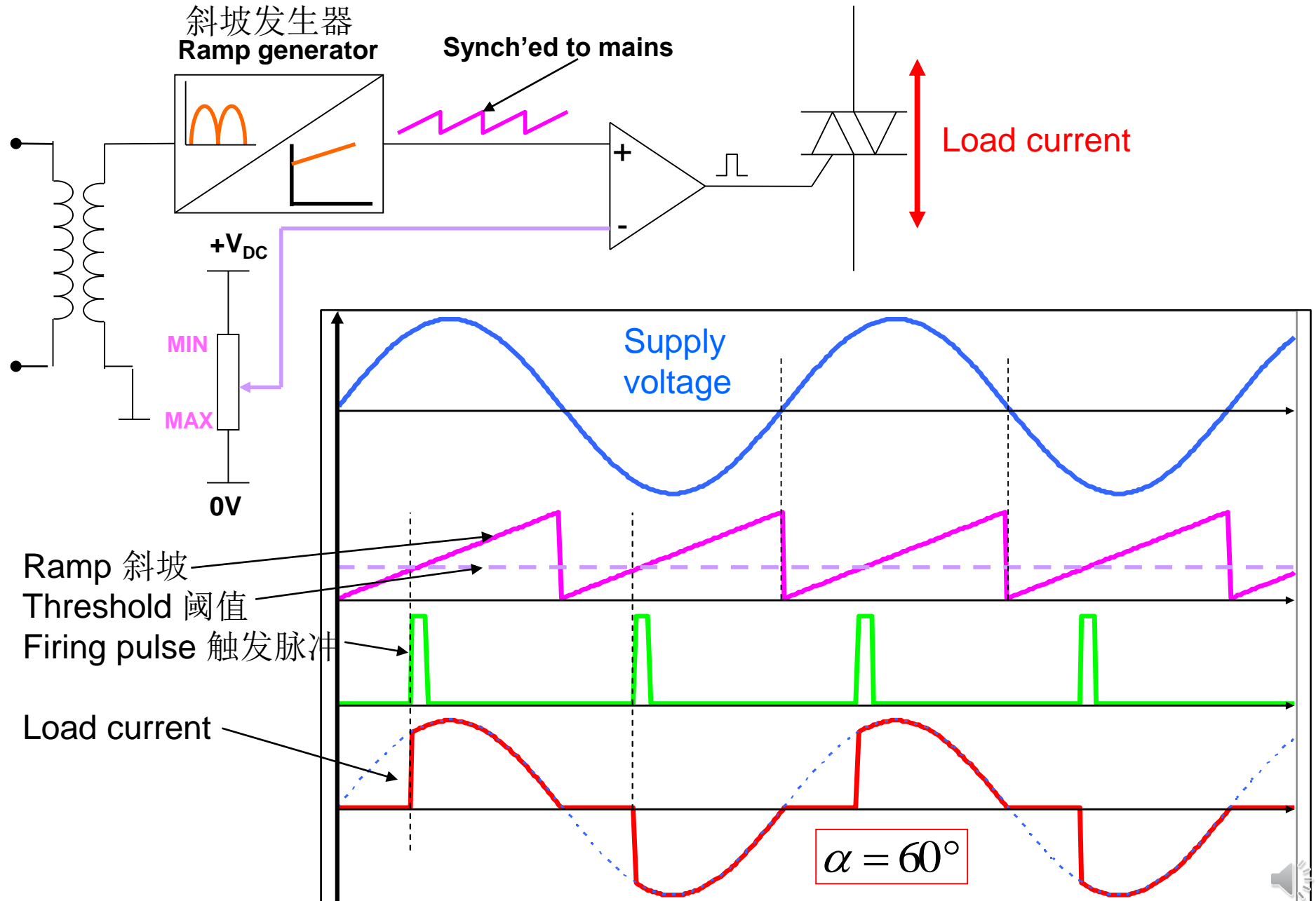
Mode I+	MT2 current+	Gate current +
Mode I-	MT2 current+	Gate current -
Mode III+	MT2 current -	Gate current +
Mode III-	MT2 current -	Gate current -



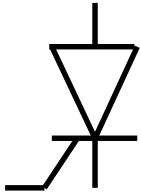
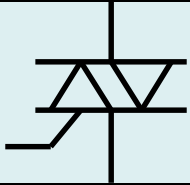
When using a TRIAC to drive an inductive load such as a motor, at each half cycle as the **current** falls to zero the line **voltage** can be quite high. The rate of change of this commutation voltage can be fast enough to interact with parasitic capacitances in the TRIAC and cause it to switch on again. This can lead to the device being permanently locked on. To circumvent this problem a **RC snubber 阻容缓冲电路** should be used across the device to **limit dV/dt** to about 1V/ μ S.



Generating the firing pulse for phase control



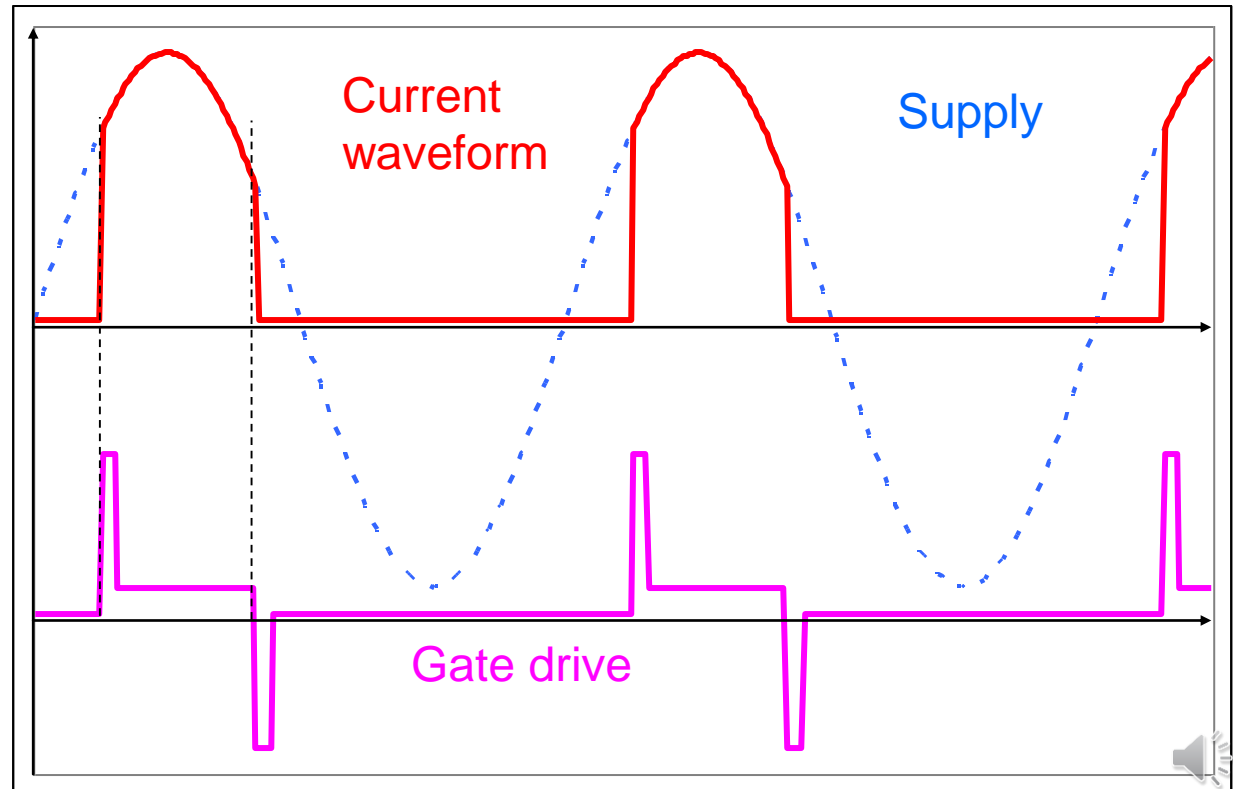
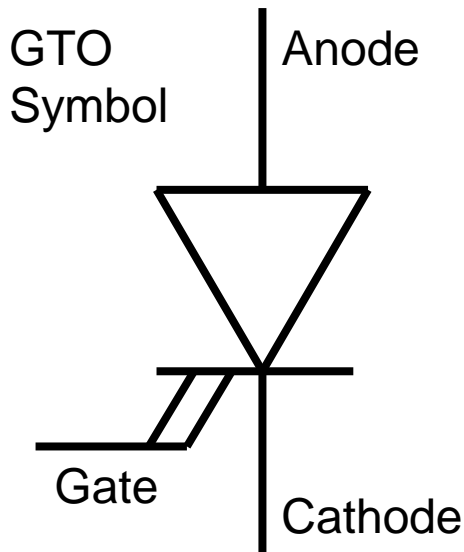
Device selection chart

Device	Symbol	Control	Power Rating	Switch speed
SCR		Partial	Very high	Slow
TRIAC		Partial	Medium	Slow
MOSFET				
IGBT				
GTO				



Gate turn-off Thyristors (GTOs)

GTOs are fully controllable SCRs. To switch on a GTO a positive pulse is applied to the gate, as in the case of the SCR. To turn off a GTO a negative pulse is applied to the gate. The turn-on of a GTO is not as reliable as an SCR (the gate sensitivity is approx 10X lower than a comparable SCR) and therefore it is usual to keep a low positive current flowing into the gate to ensure it stays on for the intended duration. GTO turn-off requires an appreciable amount of current to flow during the negative pulse and thus the drive circuitry for a GTO can be quite complex.



GTO turn-off is quite long, **several hundred μS** , and this limits the maximum switching frequency to **1KHz** or less. During turn-off the voltage must be kept low to prevent over-heating and subsequent device failure. When on, the GTO has a forward voltage drop of about **3V (3X an SCR 三倍于SCR)** and this poses additional challenges in keeping it cool.

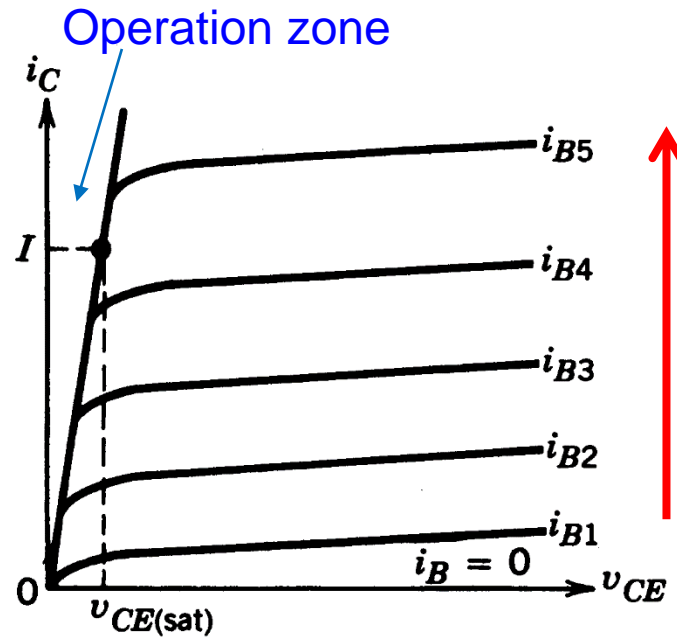
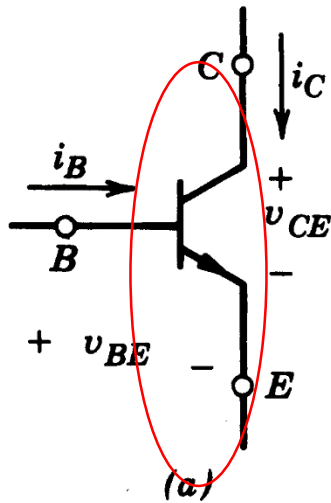
Despite the difficulties in using a GTO, they are of particular value in **high power** inverters and DC applications such as traction motor controllers where a reverse voltage (and hence natural commutation) would never occur.



6000A / 6000V GTO (Toshiba)



Bipolar Junction Transistor (BJT or GTR)



$$I_B > \frac{I_C}{h_{FE}}$$

Now GTRs are replaced by MOSFETs, IGBTs

On-State Voltage: 1~2V

Limited Voltage/Current Rating: 1400V/several hundred A

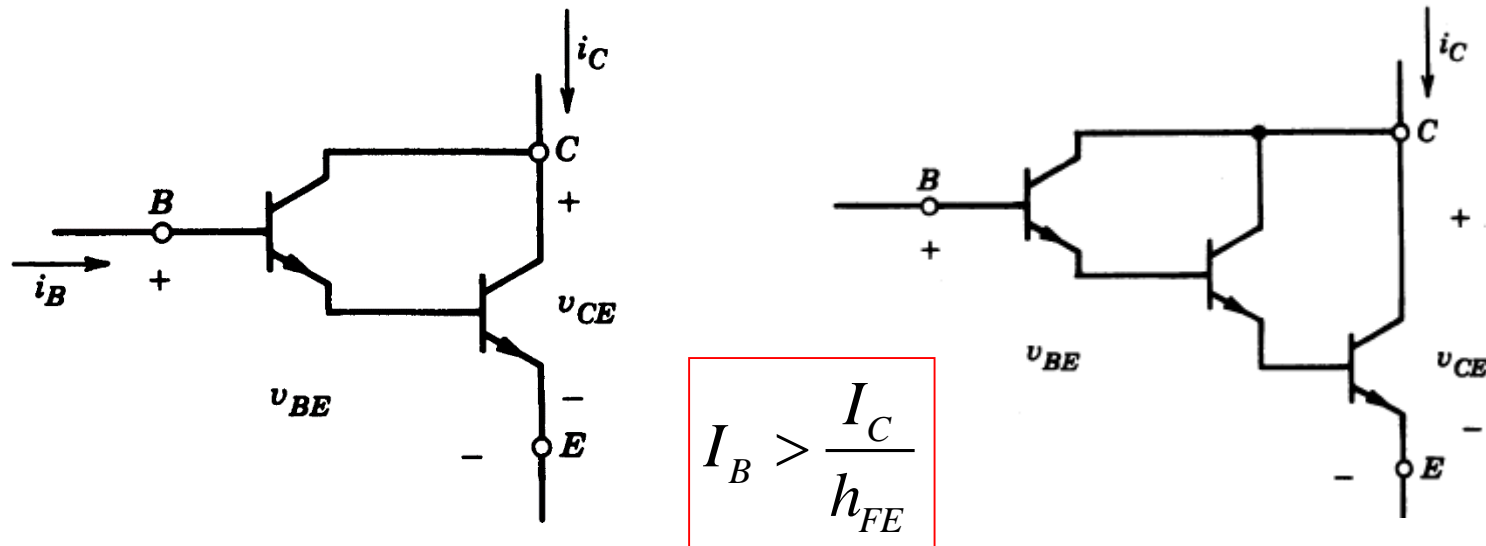
Limited Switching Frequency: several kHz

Complex current controlled drive circuit.

The letter **F** indicates that it is a forward transfer characteristic, and the letter **E** indicates it is for a common emitter configuration.



Darlington Configuration GTR

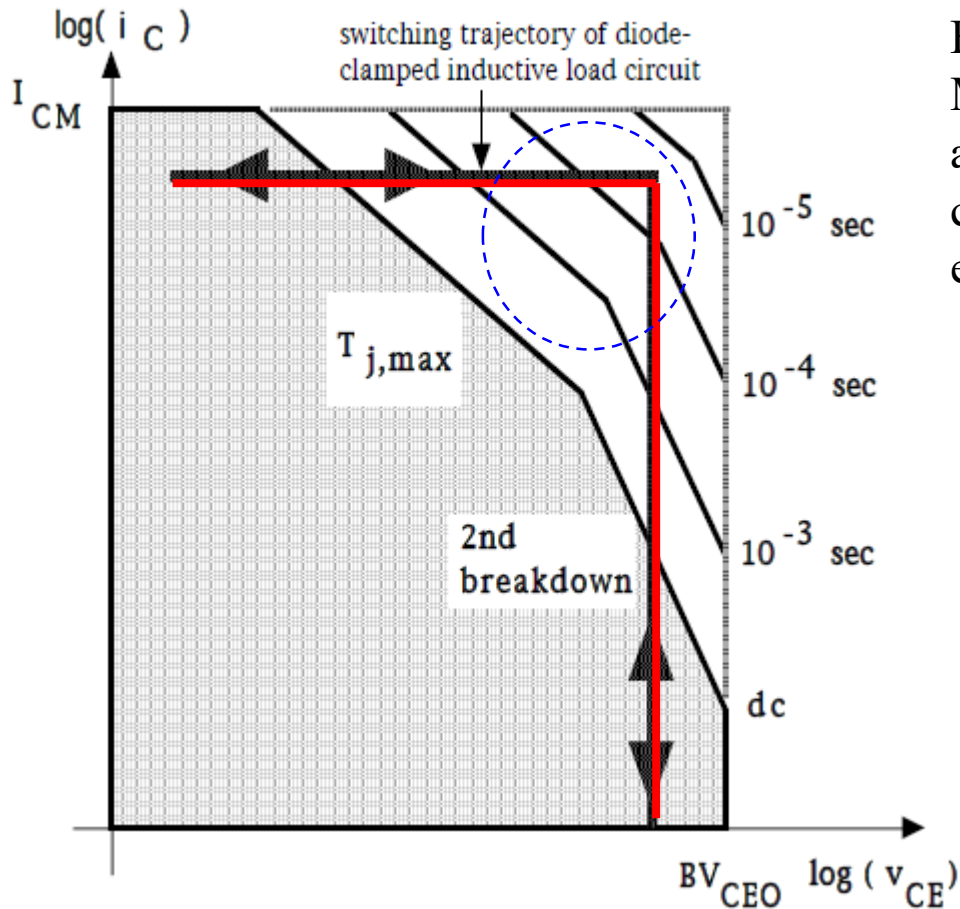


BJTs are **current controlled** (**fully-controllable**) devices, base current i_B must be supplied **continuously** to keep them in the on state. To increase the dc current gain h_{FE} and reduce the gate current loss, BJTs usually adopt Darlington configuration 达林顿结构 at cost of slightly higher $V_{CE(sat)}$ and slower switching speed. Driver circuit is also complex.



GTR Safe Operation Area

安全工作区



Forward bias 前置 safe operating area
(FBSOA)

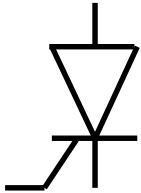
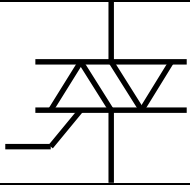
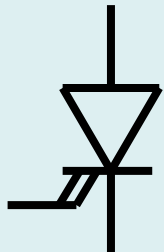
For power semiconductor devices (such as BJT, MOSFET, thyristor or IGBT), the safe operating area (SOA) is defined as the voltage and current conditions over which the device can be expected to operate without self-damage.

Snubber Circuit 缓冲电路 is needed for fully-controllable devices to shape its **transient i_C - V_{CE}** trajectory during the switching On/Off procedure.

Actually snubber is used to limit **dv/dt** and **di/dt** during switching transition.



Device selection chart

Device	Symbol	Control	Power Rating	Switch speed
SCR		Partial	Very high	Slow
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MOSFET				
IGBT				
GTO		Full	High	Slow



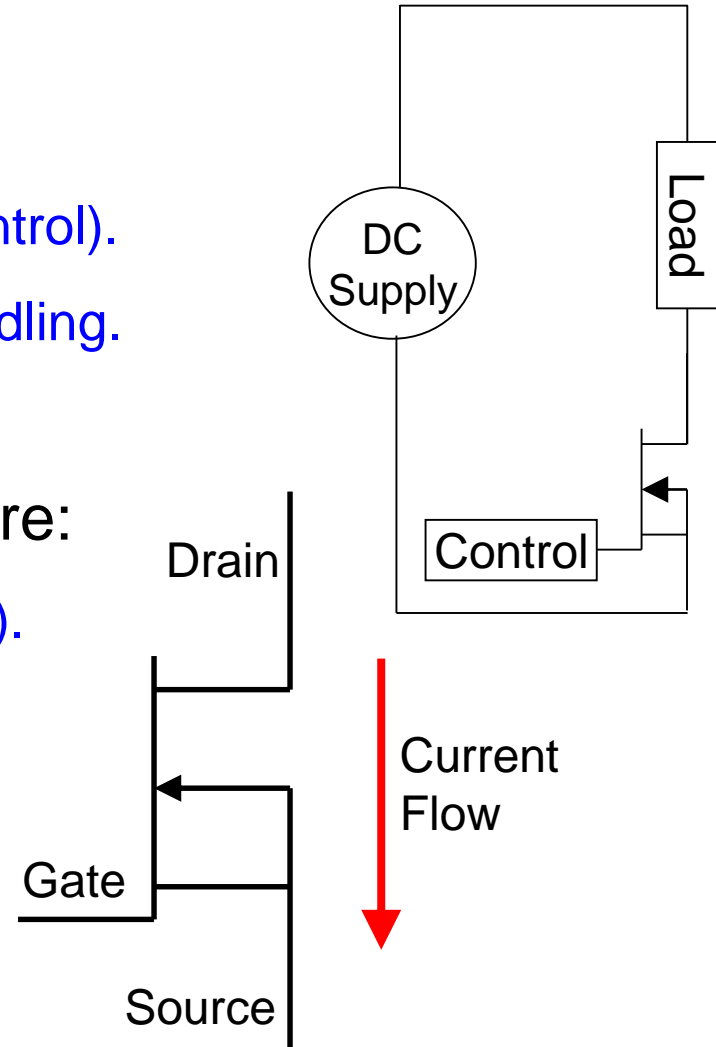
The MOSFET

MOSFETs come in many varieties 型式 and sizes 尺寸. Some of their advantages include:

- Fast switching (10's \rightarrow 100's nS).
- Huge power gain (Very simple gate control).
- Easily paralleled for higher current handling.

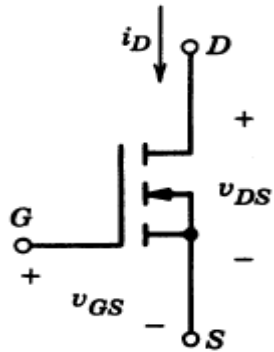
Their three main disadvantages are:

- Relatively low blocking voltage (1000V).
- Relatively low current handling (100A).
- Power dissipation can be high.

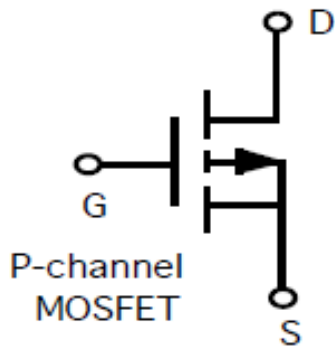


Metal Oxide Semiconductor Field-Effect Transistor (MOSFET)

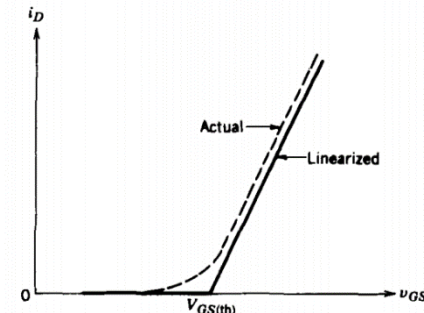
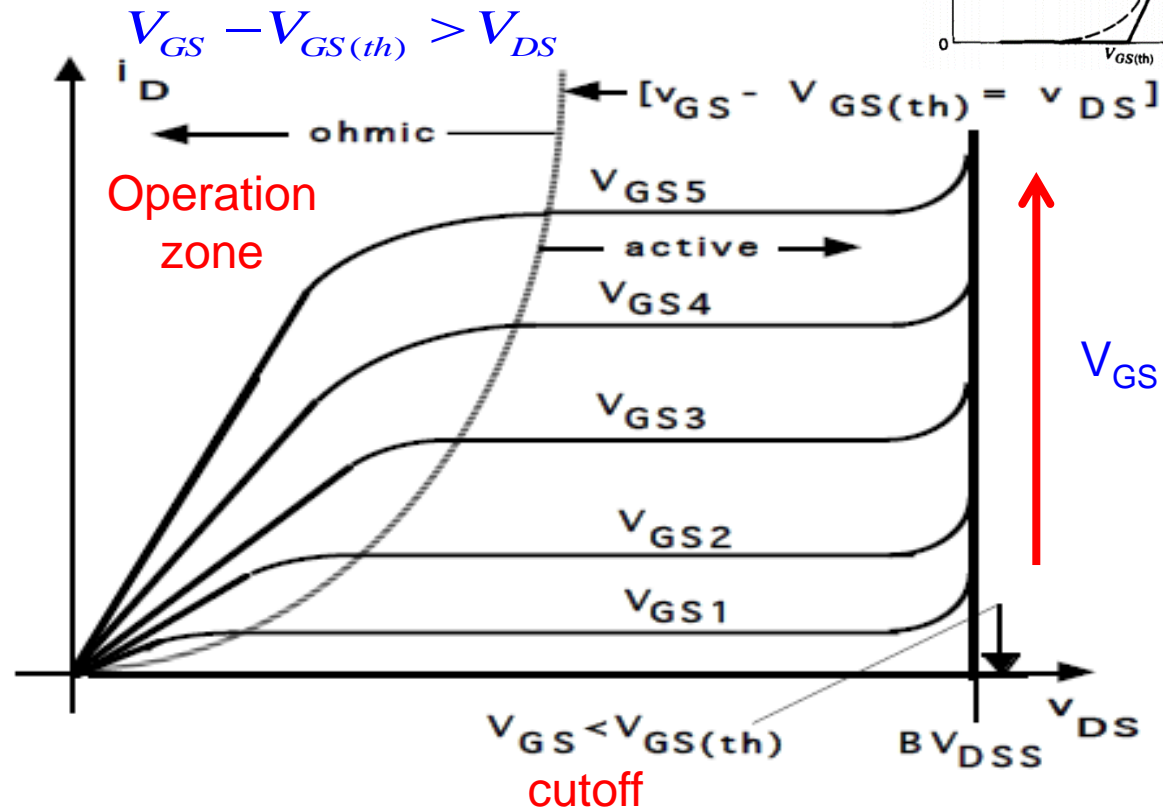
金属氧化物半导体场效应管



N-channel



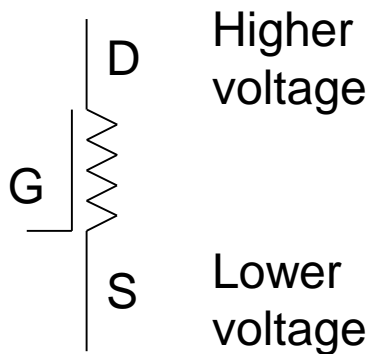
P-channel
MOSFET



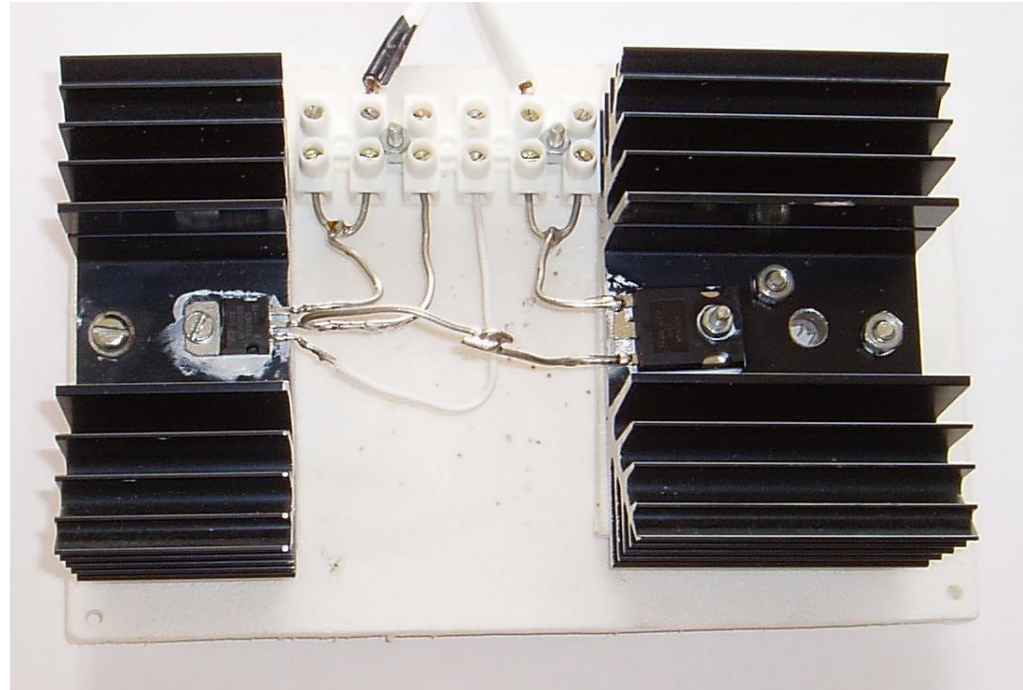
Voltage Controlled (Fully-Controllable) Device



MOSFETs can be (and often are) used for linear amplification. However, for the purposes of this course we will use them simply as **switches**. In operation, a few volts (say 5V) applied to the gate will allow a current to flow between the drain and the source. Unlike SCRs, MOSFETs can be turned on and off easily – removing the gate signal causes the MOSFET to stop conducting. **When on, the device appears as a low resistance between the drain and source.** Hence power dissipation 功率损耗 varies as I^2 .



Model – “Resistor”



The fast switching speed allows MOSFETs to be used at **operating frequencies of 100's of KHz to a few MHz**. As the MOSFET gets hot the resistance rises. This allows parallel operation because current is balanced amongst available devices automatically. There can be a significant capacitance between Drain & Source.



Characteristics of MOSFET

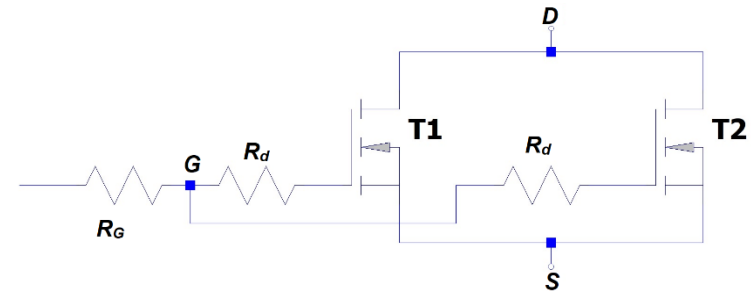
On-State Conduction Losses (Ohmic zone)

$$p_{on} = I_d^2 r_{DS(on)}$$

Good for synchronous rectification in low voltage applications.

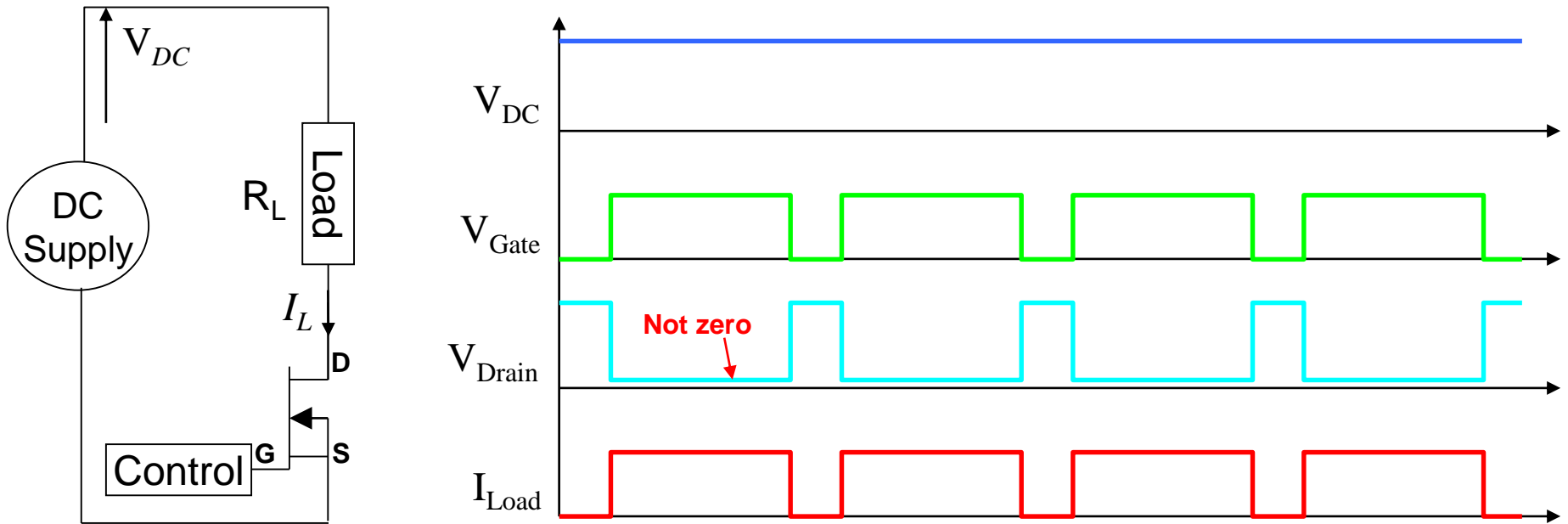
$r_{DS(on)}$ is an approximated constant resistance

- Large V_{GS} minimizes accumulation layer resistance and channel resistance
- Positive temperature coefficient: $r_{DS(on)}$ increases as temperature increases. leads to thermal stabilization effect.
- MOSFETs can be easily paralleled due to positive temperature coefficient of $r_{DS(on)}$.



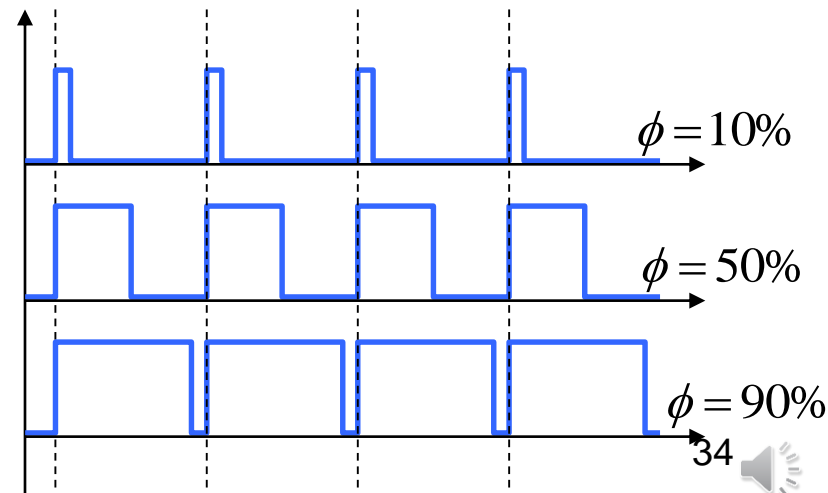
MOSFETs can be paralleled very easily, because of the positive temperature coefficient of their on-state resistance. For the same junction temperature, if $R_{DS(on)}$ of T_2 exceeds that of T_1 , then during the on state, T_1 will have a higher current and thus higher power loss compared to T_2 , since the same voltage appears across both transistors. Therefore, the junction temperature of T_1 will increase along with its on-state resistance. This will cause its share of current to decrease and, hence, there is a thermal stabilization effect.

The ease with which MOSFETs can be turned on and off makes them particularly useful in **DC applications**.



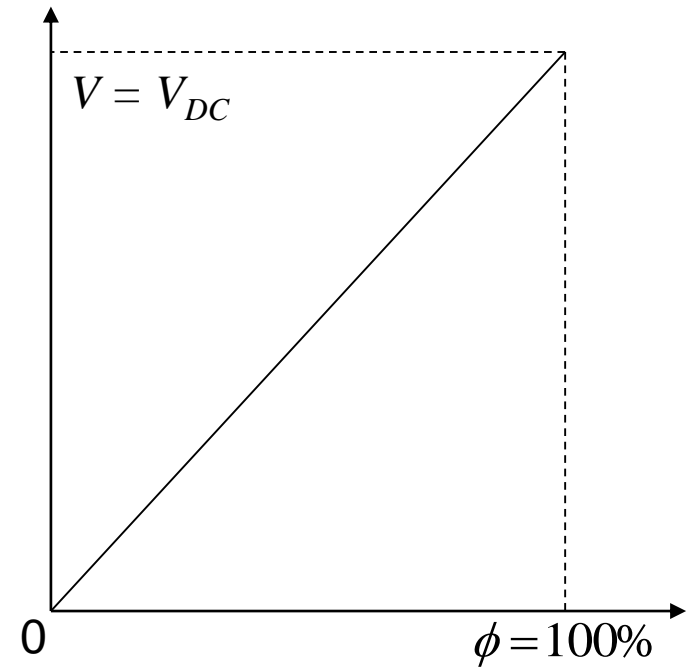
By changing the width of the gate pulse in relation to the operating frequency we can vary the width of the current pulses in the load. This is called Pulse Width Modulation.

Definition : Duty Cycle $\phi = \frac{\text{On time}}{\text{Period}}$



We can calculate the **mean DC load voltage** as a function of the duty cycle:

$$\begin{aligned} V_{\text{AVE}} &= \frac{1}{T} \int_0^T v(t) dt \\ &= \frac{1}{T} \left[\int_0^{T_{\text{ON}}} V_{\text{DC}} dt + \int_{T_{\text{ON}}}^T 0 dt \right] \\ &= \frac{V_{\text{DC}} \cdot T_{\text{ON}}}{T} = \phi \cdot V_{\text{DC}} \end{aligned}$$



If we plot this relationship we have a straight line. In other words, The average DC voltage is simply the supply voltage V_{DC} **x** the duty cycle ϕ .

Example

A heater for use with a 12V DC supply is to be designed. The heat output of the product is to be continually variable using PWM control. At 25% of the maximum setting the power output is 40W. What is the heater element resistance? (Assume a perfect switch and that there are no other power losses in the system.)

Method 1. At 25%, $P = 40\text{W}$ therefore at 100% $P = 160\text{W}$.

Power = $V^2 / R \Rightarrow R = V^2 / P = 144/160$. So the resistance = 0.9Ω .

Method 2. At 25%, the mean current is $P/V = 40/12 = 3.333\text{A}$.

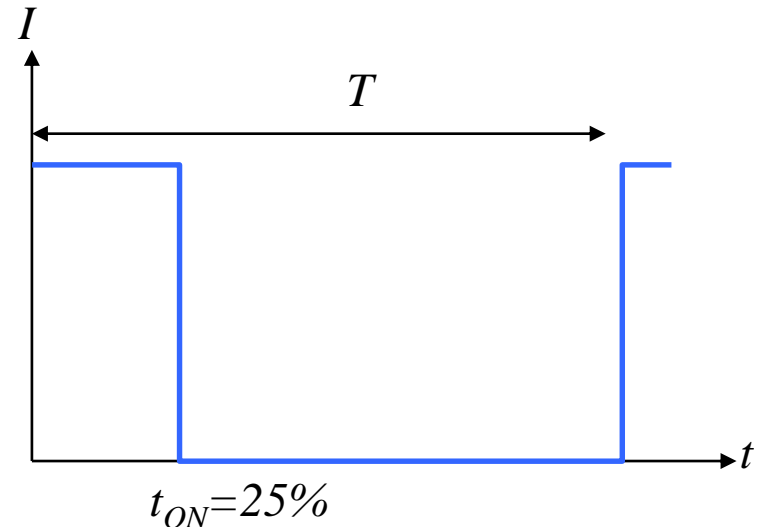
$$I_{AVE} = \frac{1}{T} \int_0^T i(t) dt. \text{ Pick } T = 1 \text{ second to make it easy:}$$

$$I_{AVE} = \int_0^1 i(t) dt = \int_0^{0.25} i(t) dt + \int_{0.25}^1 i(t) dt$$

$$\Rightarrow I_{AVE} = I_{MAX} \cdot t \Big|_{t=0}^{t=0.25}$$

$$\Rightarrow I_{MAX} = 4 \times I_{AVE} = 13.33\text{A}$$

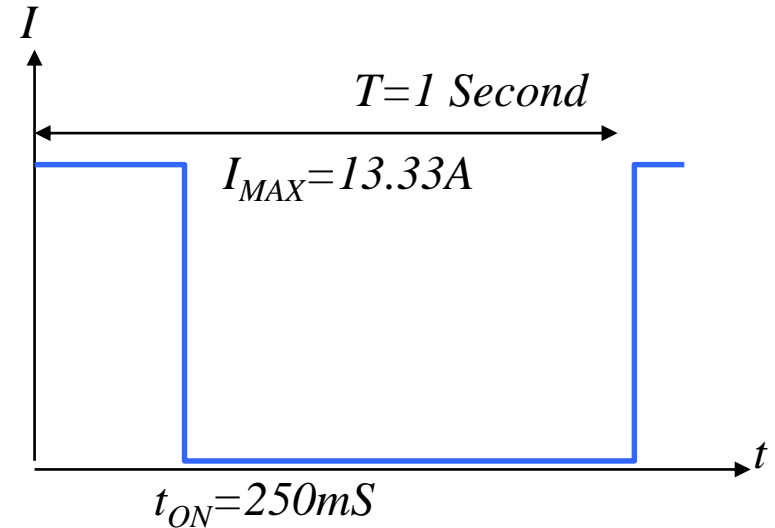
$$R = \frac{V}{I} = \frac{12}{13.33} = 0.9\Omega$$



Proof of RMS being the heating value

At 25%, the RMS value of the waveform is:

$$\begin{aligned} I_{RMS} &= \sqrt{\frac{1}{T} \int_0^T i(t)^2 dt} \\ &= \sqrt{\int_0^{0.25} I_{MAX}^2 dt + \int_{0.25}^1 0 dt} \\ &= \sqrt{I_{MAX}^2 \cdot t \Big|_0^{0.25}} \\ &= \sqrt{0.25 I_{MAX}^2} = \sqrt{\frac{178}{4}} = 6.66 A \end{aligned}$$



We also know $P = I^2 R$, $R = 0.9 \Omega$, $P = 40 W$.

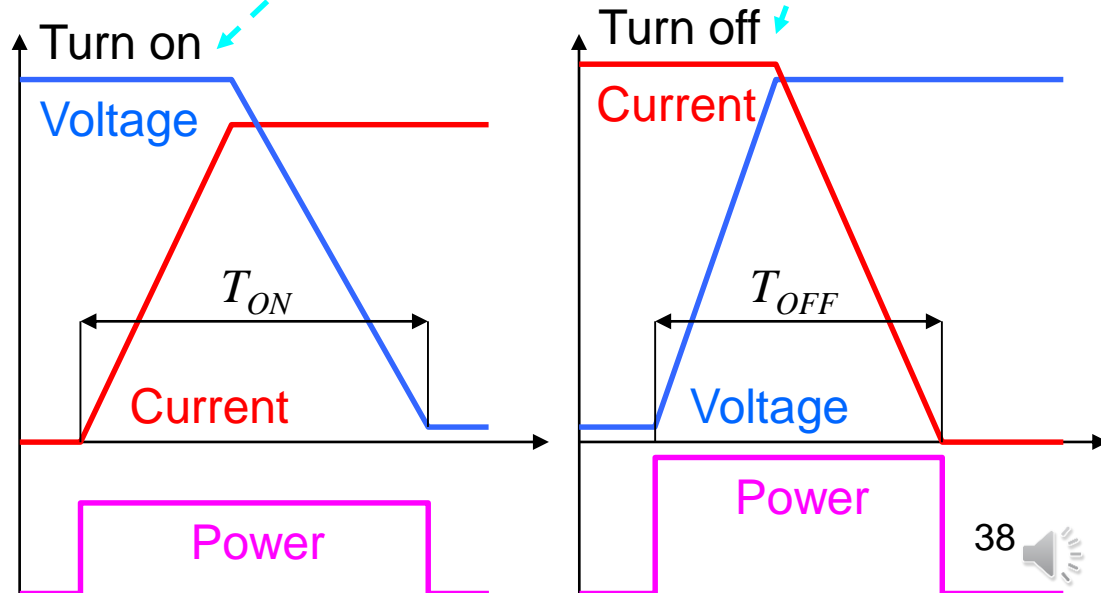
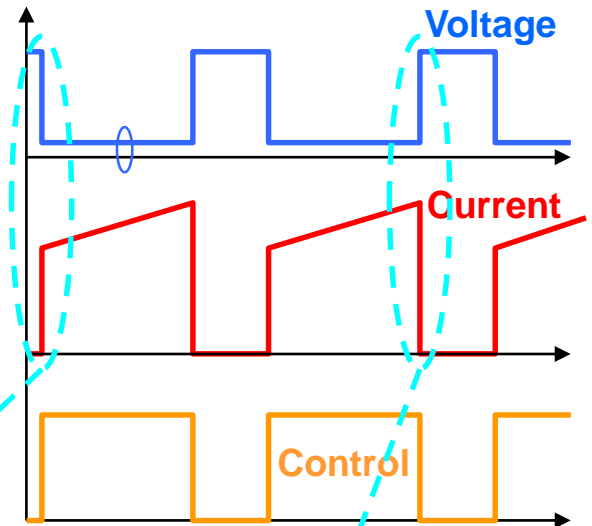
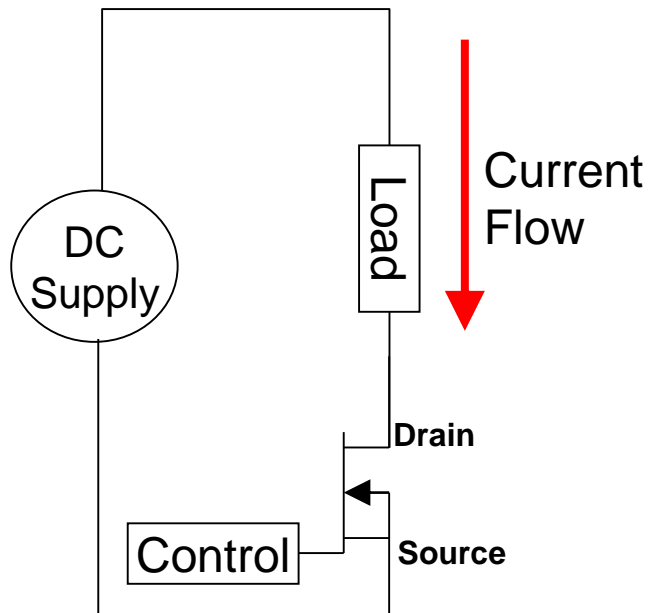
$$\Rightarrow I^2 = \frac{P}{R} = \frac{40}{0.9} = 44.4$$

$$\Rightarrow I_{RMS} = 6.66 \text{ Amps}$$

The MOSFET switching loss needs to be carefully considered: there are two switching edges for each cycle of the periodic waveform.

At switch-on, V_{DS} falls to a low value determined by I_{ON} and the equivalent resistance $R_{DS\ ON}$.

When driving inductive loads, the current through the switch will not usually be a constant value during the cycle.



Power dissipation in MOSFETs

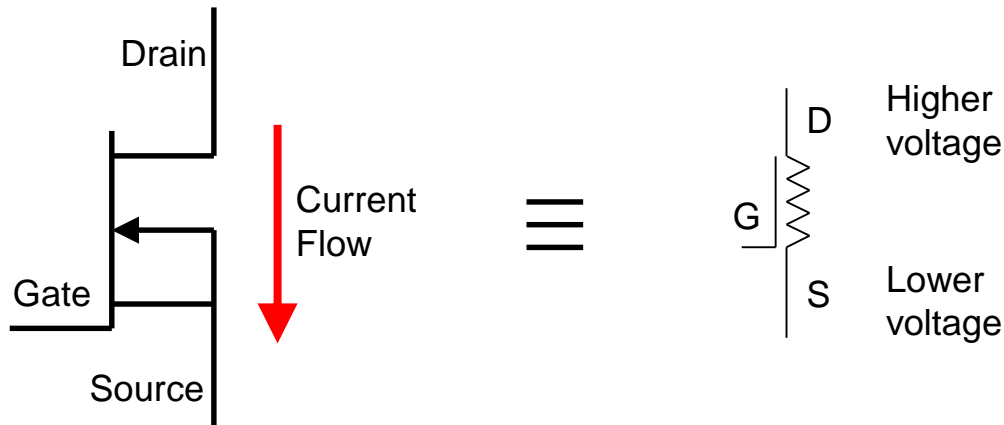
There are two main power loss mechanisms in a MOSFET:

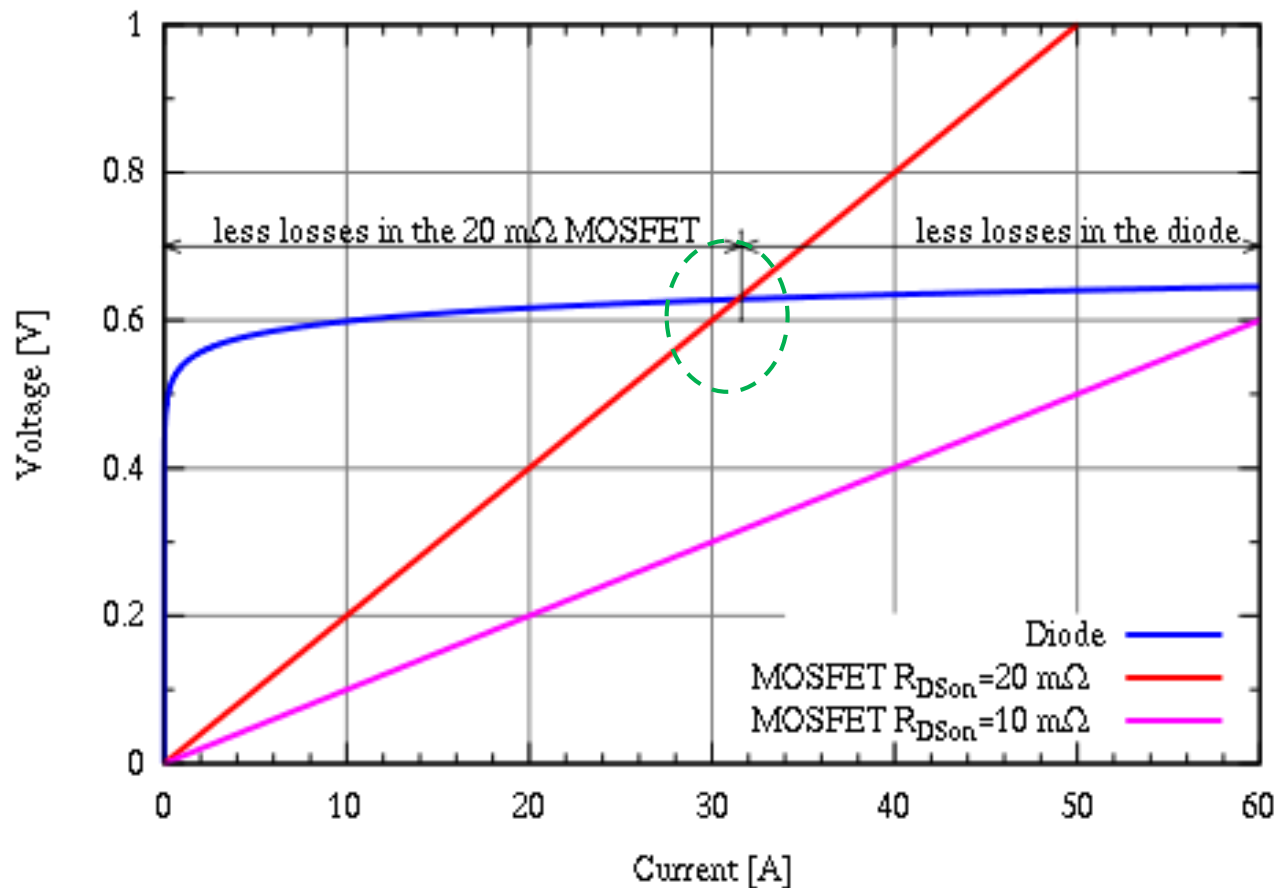
$$\text{Power loss} = \text{Conduction loss} + \text{Switching loss}$$

The conduction loss is the I^2R loss due to the load current flowing through the equivalent resistance between the drain and source. This depends directly on the duty cycle the switch is operating at.

Calculate the RMS load current to include the duty cycle. (See example in a few pages time.)

$$P_{\text{Conduction}} = I_{\text{Load RMS}}^2 R_{\text{DS ON}}$$

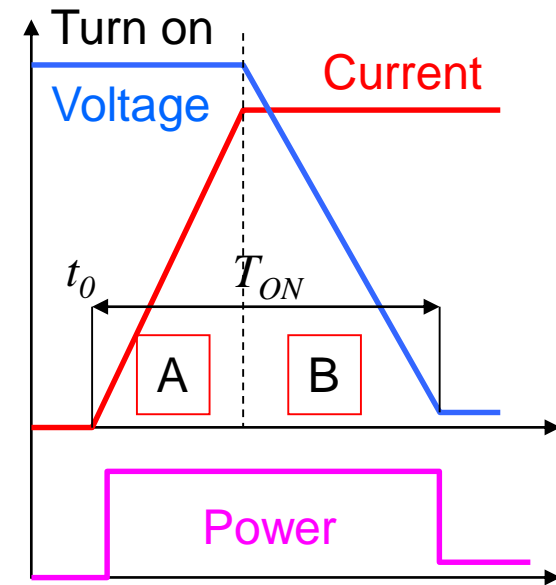




MOSFET is not suitable for high power applications !!!

- In the low current flow range, the voltage drop across the MOSFET is less than the almost constant voltage drop across diode/GTR/SCR, the conduction power loss of MOSFET is less than that of Diode/SCR/....
- Voltage and Current ratings are only up to 1kV and several hundred Amperes.

During turn-on, at switching event t_0 the current through the FET starts to rise towards the “on” value. During this period (A) the voltage across the FET ($V_{DS(Off)}$) is essentially constant. Once the current reaches the “on” value, the voltage across the switch then starts to fall (B) to the conduction value, determined by $R_{DS ON}$. When this value is reached the FET is in the “on” state. The time for this sequence of events to happen, T_{ON} , is specified in the device’s data sheet. A typical value is 25nS.



During ‘A’, V_{DS} is constant at $V_{DS(Off)}$ and $I_{Ave} = \frac{1}{2} I_{ON} \Rightarrow E_A = \frac{1}{2} I_{ON} V_{DS(Off)} T_A$.

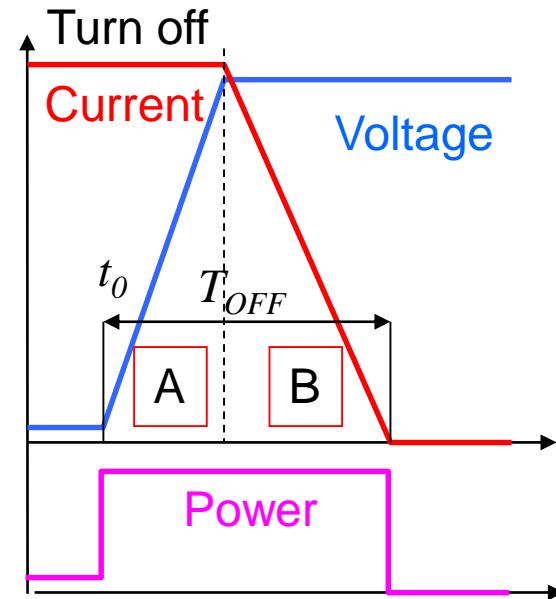
During ‘B’ I_{ON} is constant and $V_{DS AVE} = \frac{1}{2} V_{DS(Off)} \Rightarrow E_B = \frac{1}{2} I_{ON} V_{DS(Off)} T_B$.

The energy during the switching event is thus $E_A + E_B = \frac{1}{2} I_{ON} V_{DS(Off)} T_{ON}$.

The total power per second for **switch-on** is given by:

$$\text{Total "on" power} = \frac{1}{2} (F_{SW} \cdot T_{ON} \cdot I_{ON} \cdot V_{DS(Off)})$$

During turn-off, at switching event t_0 the voltage across the FET starts to rise towards the “off” value $V_{DS(Off)}$. During this period (A) the current through the FET (I_{OFF}) is essentially constant at the value just before the turn-off event started. Note this is unlikely to be the same as I_{ON} . Once the voltage reaches the “off” value, the current through the switch then starts to fall (B) to zero. When this value is reached the FET is in the “off” state. The time for this sequence of events to happen, T_{OFF} , is specified in the device’s data sheet. A typical value is 30nS.



During ‘A’ I_{OFF} is constant and $V_{DS\,AVE} = \frac{1}{2}V_{DS(Off)} \Rightarrow E_A = \frac{1}{2}I_{OFF}V_{DS(Off)}T_A$.

During ‘B’, V_{DS} is constant at $V_{DS(Off)}$ and $I_{Ave} = \frac{1}{2}I_{OFF} \Rightarrow E_B = \frac{1}{2}I_{OFF}V_{DS(Off)}T_B$.

The energy during the switching event is thus $E_A + E_B = \frac{1}{2}I_{OFF}V_{DS(Off)}T_{OFF}$.

The total power per second for **switch-off** is given by:

$$\text{Total "off" power} = \frac{1}{2} (F_{SW} \cdot T_{OFF} \cdot I_{OFF} \cdot V_{DS(Off)})$$

The total switching loss for the MOSFET is the sum of the ON and OFF losses:

$$P_{\text{Switching}} = \left(\frac{F_{\text{SW}} \cdot V_{\text{DS(Off)}}}{2} \right) (T_{\text{ON}} \cdot I_{\text{ON}} + T_{\text{OFF}} \cdot I_{\text{OFF}})$$

We can now calculate the total power loss in the MOSFET due to the conduction and switching losses.

Remember $P_{\text{Conduction}} = I_{\text{Load RMS}}^2 R_{\text{DS ON}}$

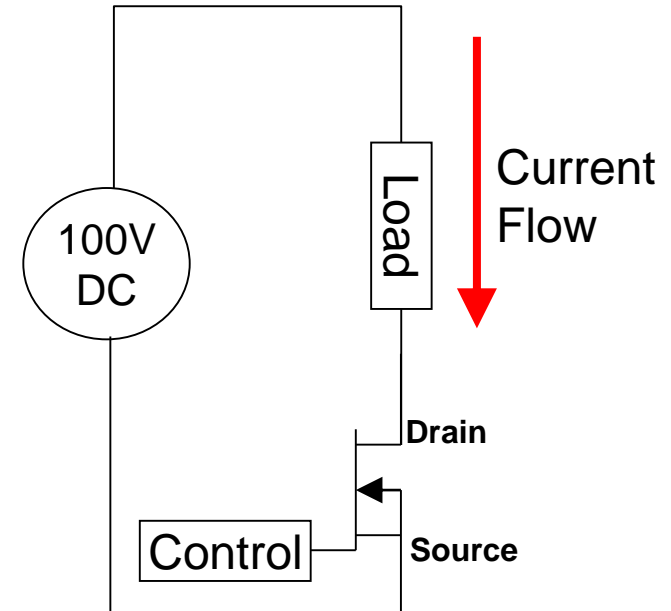
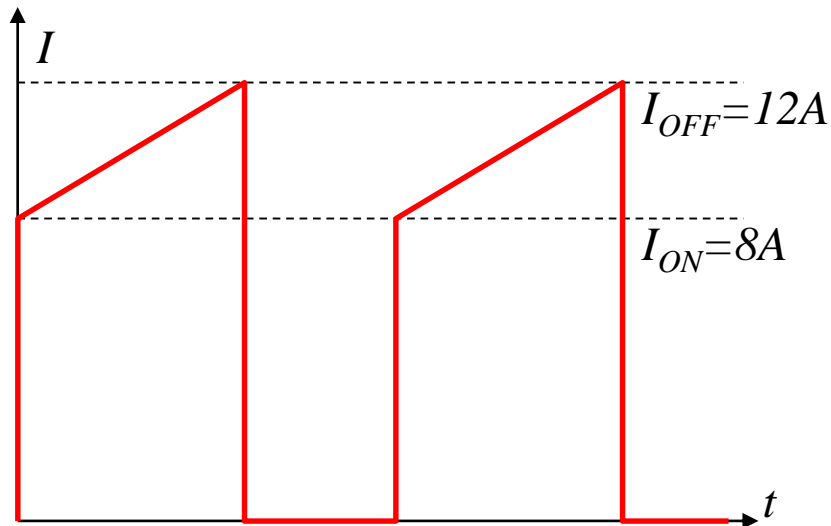
$$P_{\text{Total}} = I_{\text{Load RMS}}^2 R_{\text{DS ON}} + \left(\frac{F_{\text{SW}} \cdot V_{\text{DS(Off)}}}{2} \right) (T_{\text{ON}} \cdot I_{\text{ON}} + T_{\text{OFF}} \cdot I_{\text{OFF}})$$

From data sheet

Example

Using the following data, determine the power loss in the MOSFET.

I_{ON}	8A
I_{OFF}	12A
PWM Duty	60%
PWM Freq	50KHz
Supply	100V DC
R_{ON}	50m Ω
T_{ON}	25nS
T_{OFF}	30nS



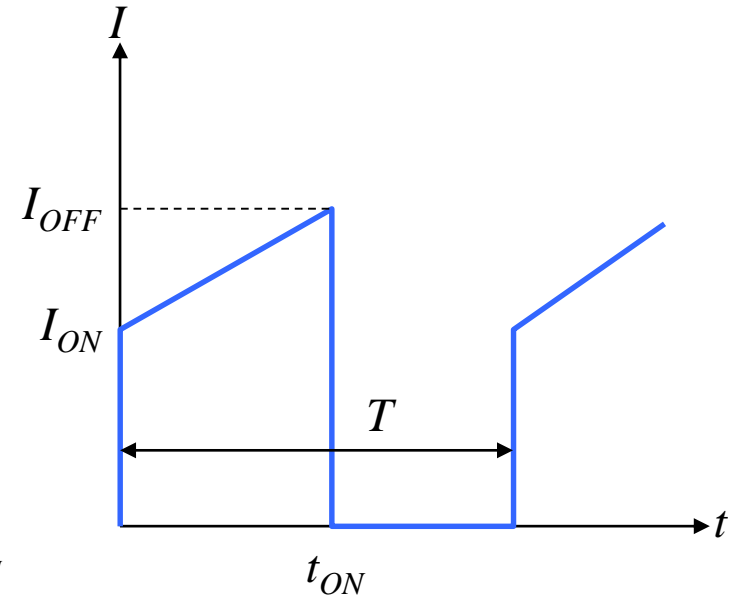
PWM period = 20 μ S
On time = 12 μ S
Off time = 8 μ S

Step 1 - Calculation of RMS current for the conduction loss.

Current function : $i(t) = I_{ON} + \frac{(I_{OFF} - I_{ON})t}{t_{ON}}; 0 < t \leq t_{ON}$

$$(i(t))^2 = I_{ON}^2 + \frac{2I_{ON}(I_{OFF} - I_{ON})t}{t_{ON}} + \frac{(I_{OFF} - I_{ON})^2 t^2}{t_{ON}^2}$$

$$\begin{aligned} I_{RMS}^2 &= \frac{1}{T} \int_0^{t_{ON}} \left[I_{ON}^2 + \frac{2I_{ON}(I_{OFF} - I_{ON})t}{t_{ON}} + \frac{(I_{OFF} - I_{ON})^2 t^2}{t_{ON}^2} \right] dt \\ &= \frac{1}{T} \left[I_{ON}^2 \cdot t + \frac{I_{ON}(I_{OFF} - I_{ON})t^2}{t_{ON}} + \frac{1}{3} \cdot \frac{(I_{OFF} - I_{ON})^2 t^3}{t_{ON}^2} \right]_0^{t_{ON}} \\ &= \frac{t_{ON}}{T} \left[I_{ON}^2 + I_{OFF} \cdot I_{ON} + \frac{1}{3} \cdot (I_{OFF}^2 - 2I_{OFF} \cdot I_{ON} + I_{ON}^2) \right] \\ &= \frac{\phi}{3} [I_{OFF}^2 + I_{OFF} \cdot I_{ON} + I_{ON}^2] \end{aligned}$$



For $I_{ON} = 8A$, $I_{OFF} = 12A$ and $\phi = 60\%$, $I_{RMS} = 7.8A$

$$P_{Conduction} = I_{Load RMS}^2 R_{DS ON}$$

$$\Rightarrow P_{Conduction} = 7.8^2 \times 50m\Omega = 3.042W$$

Step 2 – calculation of switching loss.

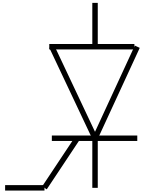
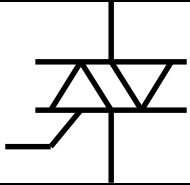

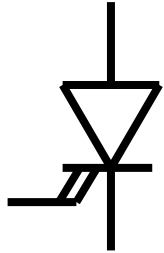
$$\begin{aligned} P_{\text{Switching}} &= \left(\frac{F_{\text{SW}} \cdot V_{\text{DS(Off)}}}{2} \right) (T_{\text{ON}} \cdot I_{\text{ON}} + T_{\text{OFF}} \cdot I_{\text{OFF}}) \\ &= \left(\frac{50\text{KHz} \times 100\text{V}}{2} \right) (25\text{nS} \times 8\text{A} + 30\text{nS} \times 12\text{A}) \\ &= 1.4\text{W} \end{aligned}$$

Step 3 – calculate total power loss

Power loss = Conduction loss + Switching loss

$$= 3.04\text{W} + 1.4\text{W} = 4.44\text{W}$$

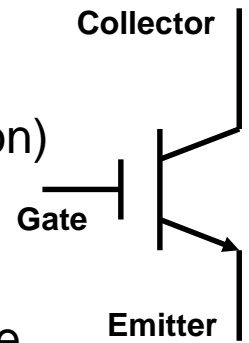
Device selection chart

Device	Symbol	Control	Power Rating	Switch speed
SCR		Partial	Very high	Slow
TRIAC & SSR		Partial	Medium	Slow
MOSFET		Full	Medium	Fast
IGBT				
GTO		Full	High	Slow



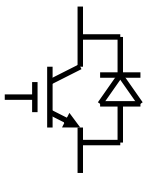
Insulated Gate Bipolar Transistor (IGBT)

IGBTs are medium to high power switches that can operate very quickly with high current and a low “on” voltage. The latest devices (3rd generation) are extremely robust and tolerant of overloads. The IGBT gate drive requirements are very similar to those of a MOSFET.



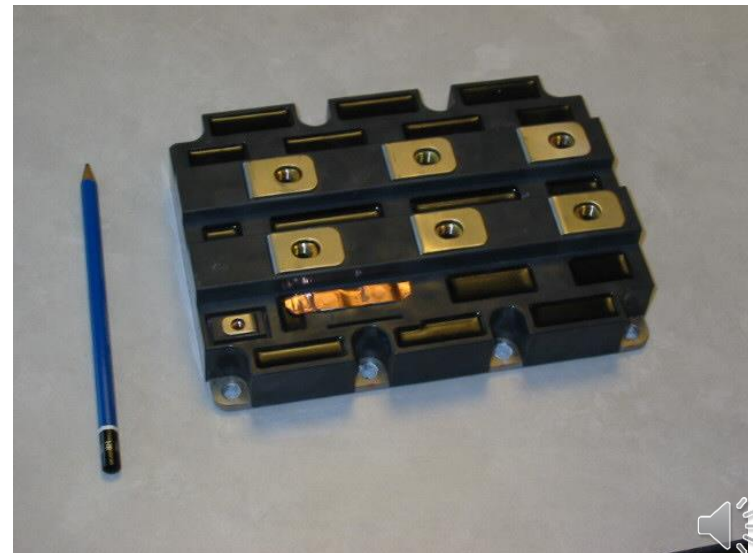
Rather conveniently the typical application reverse blocking voltage range for the IGBT starts more or less where the MOSFET starts to struggle. IGBTs are rarely used in application where the reverse voltage is less than 600V.

IGBTs can only conduct current in one direction. If current has to flow in the opposite direction then an external diode must be used in inverse parallel with the device. This is sometimes called a “flyback” 反激 diode or a “free-wheeling” diode 续流二极管 (from its first use in television sets).

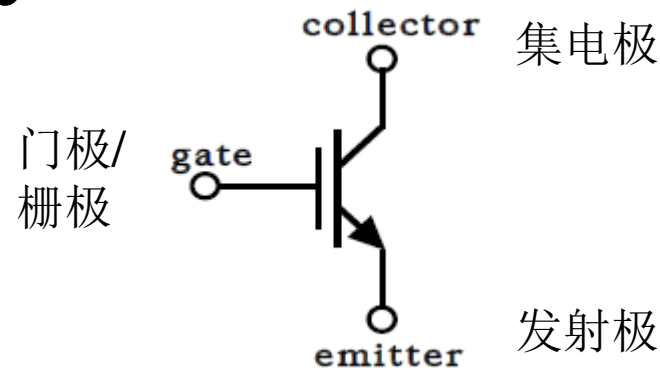
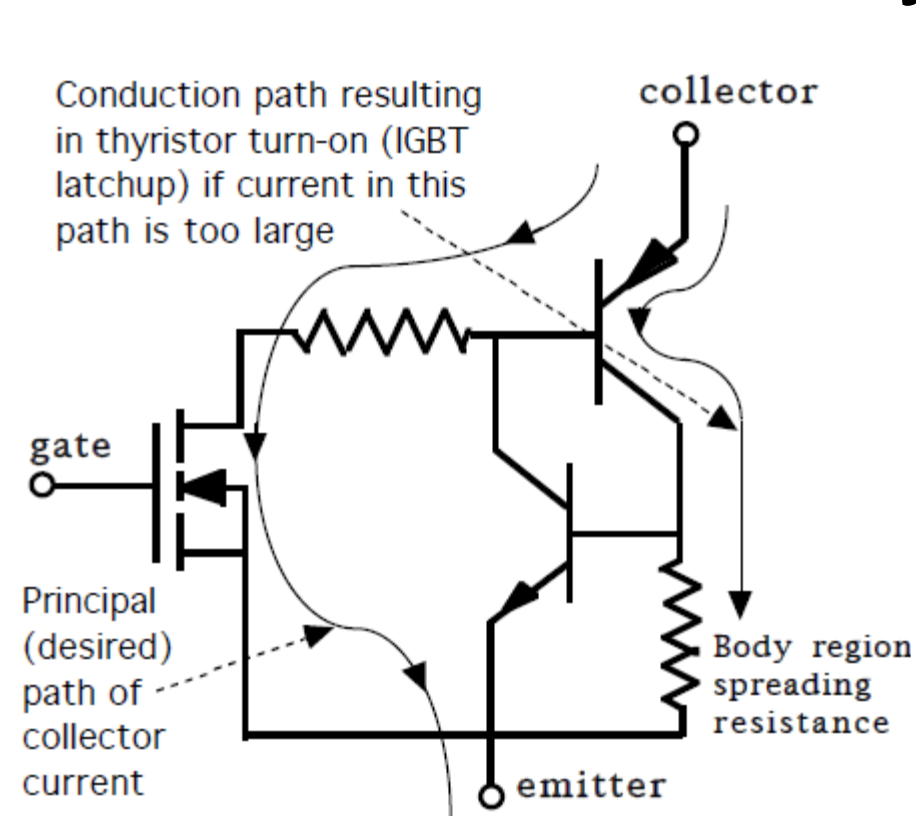


The Toyota Prius hybrid electric car uses IGBTs in its regenerative inverter drive.

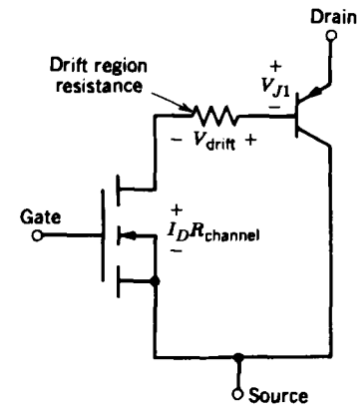
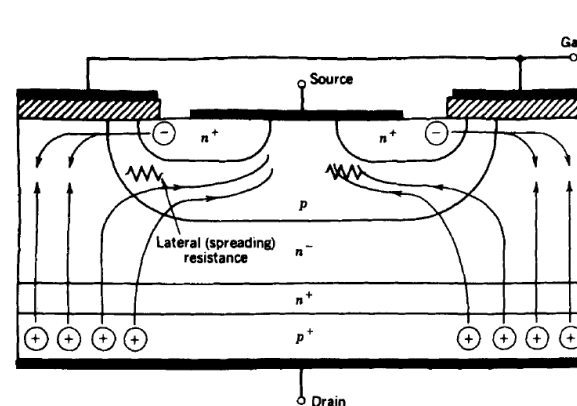
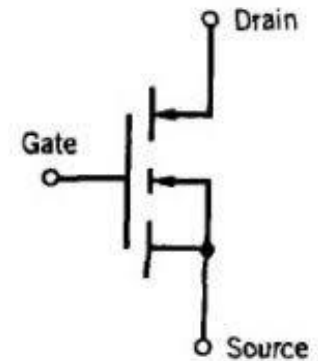
[The car has been declared a hazard to the visually impaired because it is so quiet.]



IGBT - Hybrid Switch



N-channel IGBT



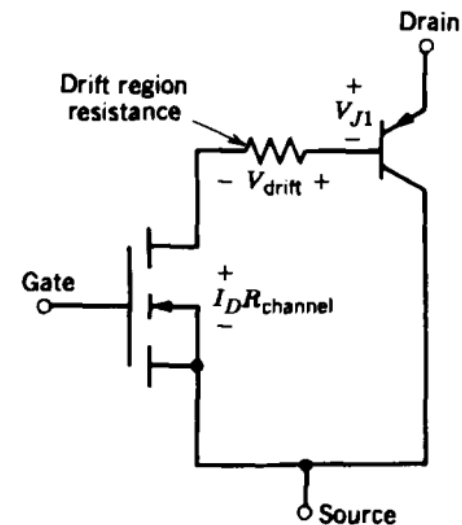
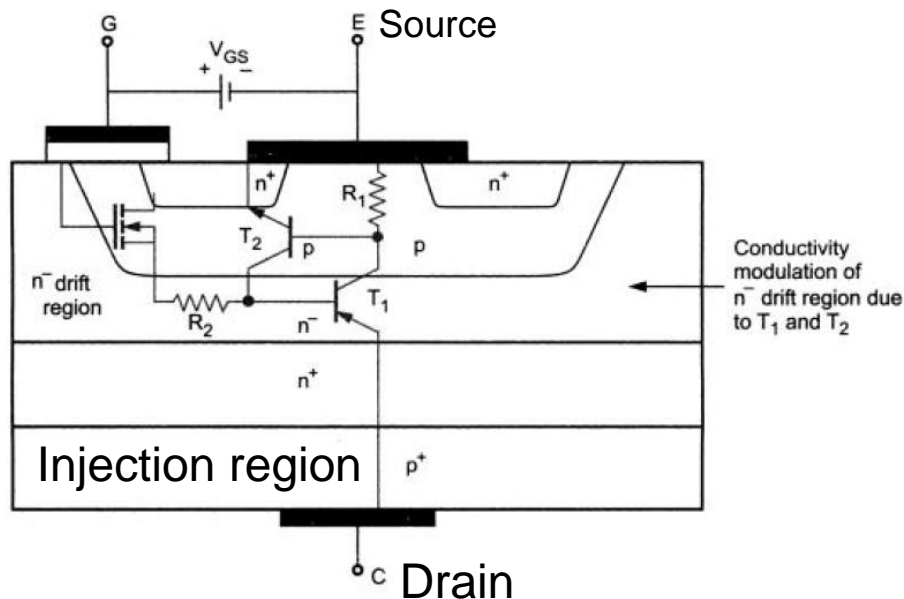
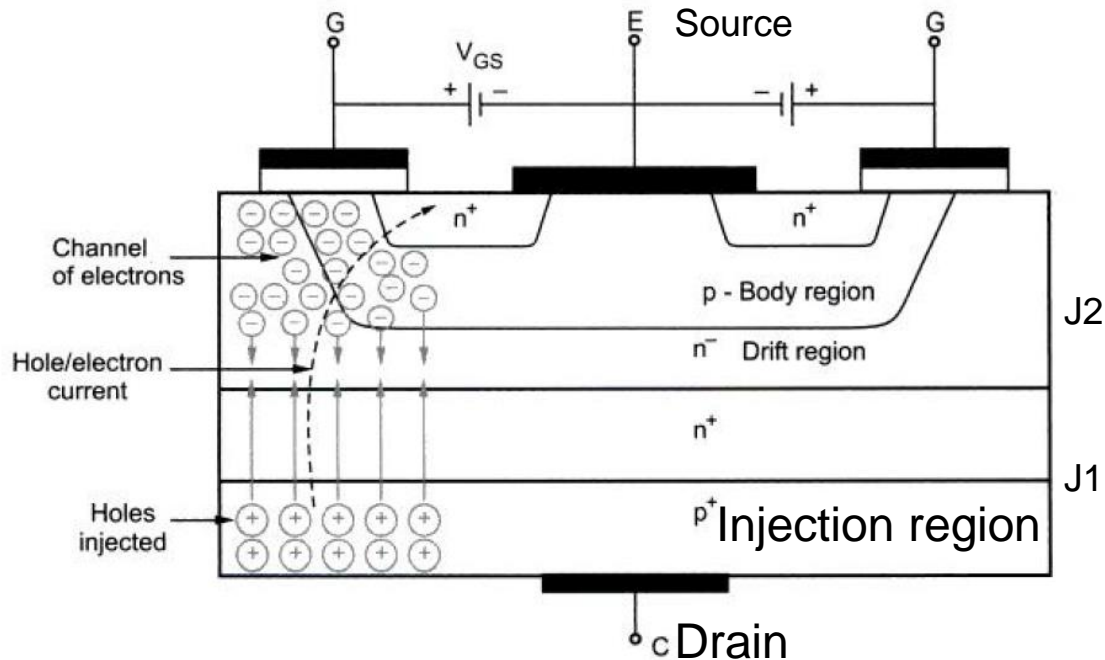
N-channel IGBT Equivalent Circuit

$$V_{DS(on)} = V_{J1} + V_{drift} + I_D R_{channel}$$

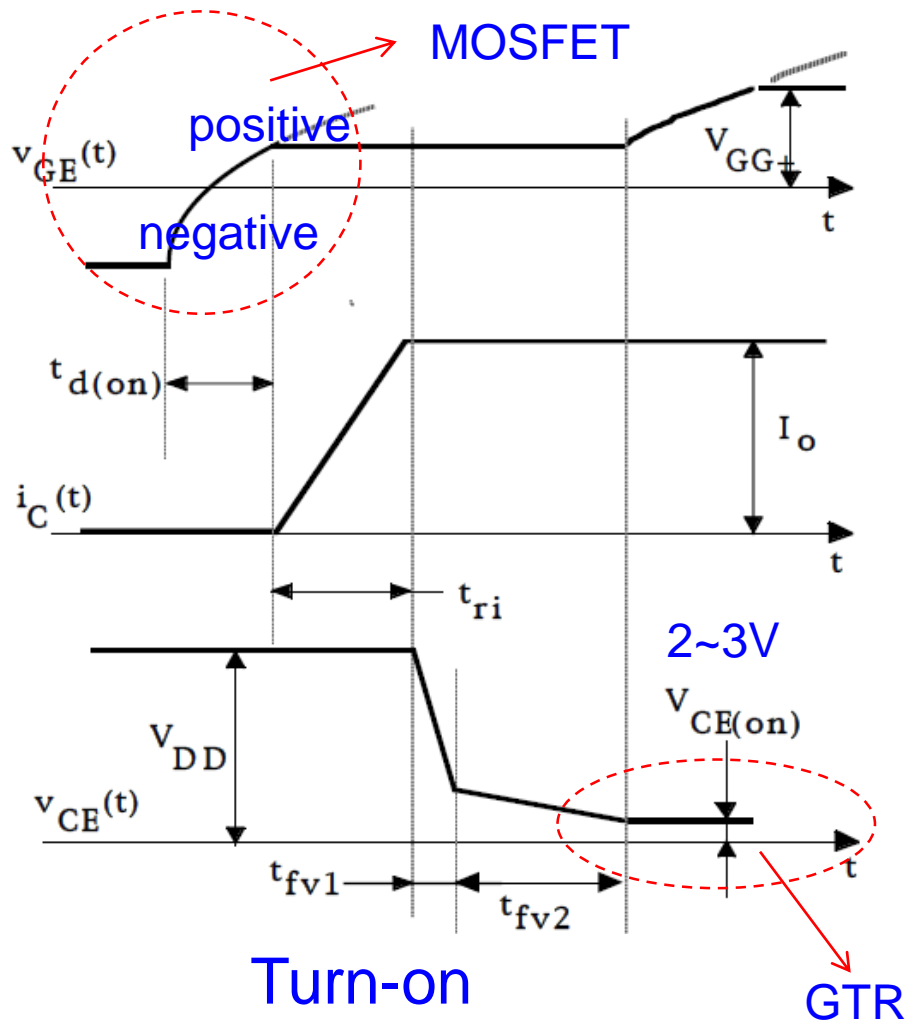
- Gate Voltage V_{GE} Controlled Device like MOSFET
- Turn-on V_{CE} like GTR: 2~3V



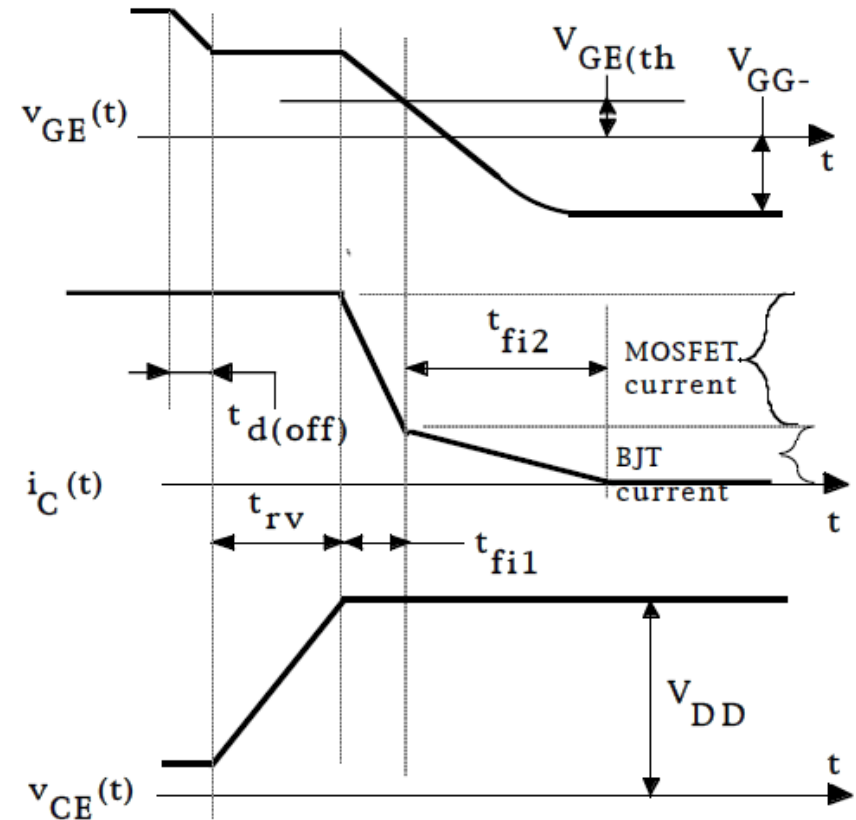
Internal structure of IGBT



Turn On/Off of IGBT



$$t_{on} = t_{d(on)} + t_{ri} + t_{fv}$$



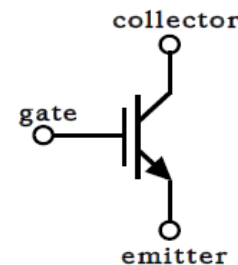
Turn-off

$$t_{off} = t_{d(off)} + t_{rv} + t_{fi1} + t_{fi2}$$

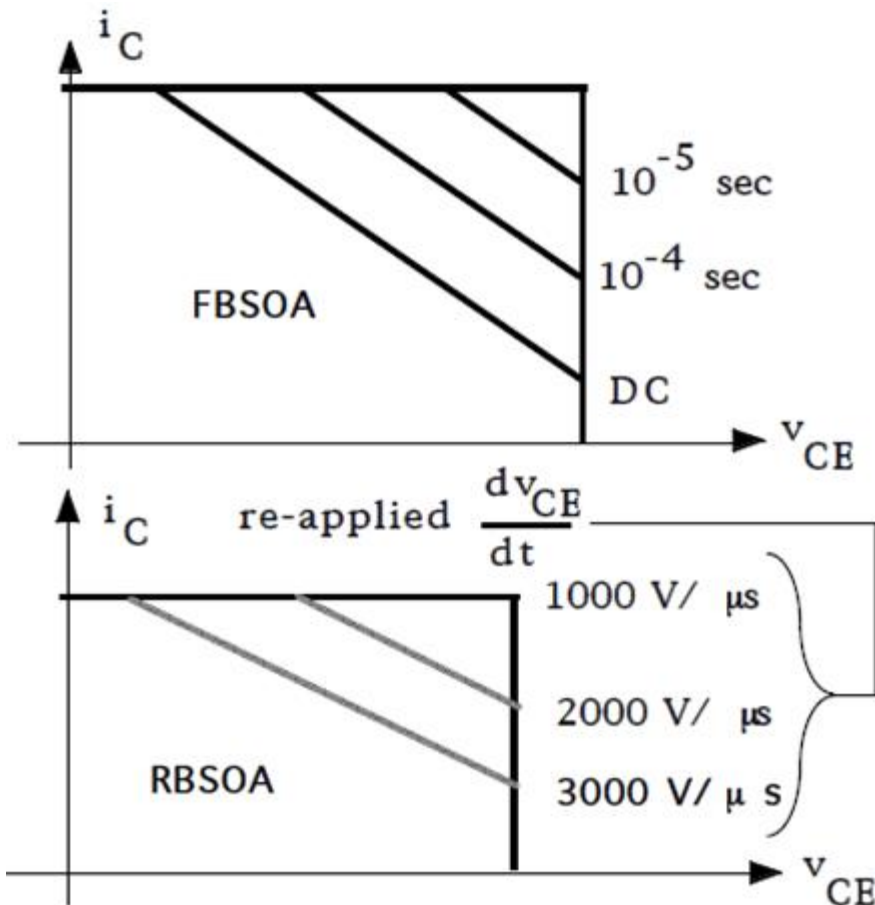
t_{rv} is voltage rise time.

t_{fi1} is MOSFET current fall time.

t_{fi2} is BJT current fall time.



SOA for IGBT



Voltage/current ratings of IGBT, MOSFET and BJT in SOA and datasheets are peak values, not **average** or **rms** value

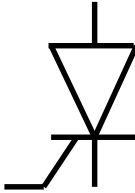
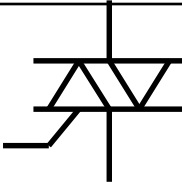
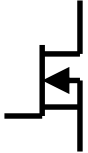
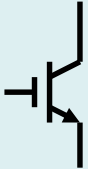
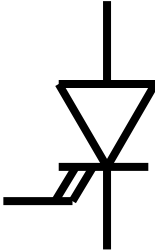
Snubber circuits are needed to shape **i-v** curve for IGBT during switching transition.

- 1 Forward Biased Safe operating area(FBSOA)-during turn on
2. Reverse Biased Safe Operating Area(RBSOA)-during turn off

***Too large a value of dV_{CE}/dt is during turn-off will cause latchup the IGBT**



Device selection chart

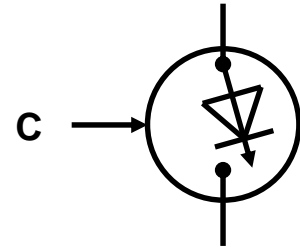
Device	Symbol	Control	Power Rating	Commutation	Supply Type	Switch speed
SCR		Partial	Very high	Natural	AC	Slow
TRIAC & SSR		Partial	Medium	Natural	AC	Slow
MOSFET		Full	Medium	Self	DC	Fast
IGBT		Full	High	Self	DC	Medium
GTO		Full	High	Self	AC/DC	Slow



Commutation 换流

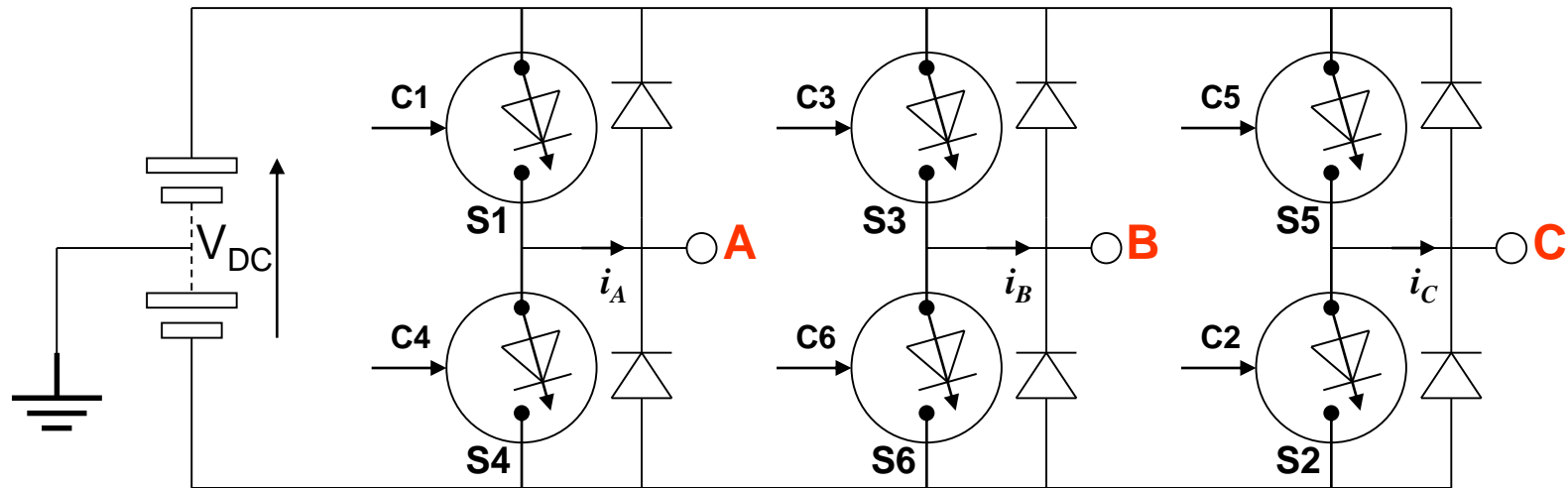
Often we want to specify in circuit schematics the operation of the device, rather than the device itself. A good example is the *self-commutated* 自换流 semiconductor switch. Devices which are **fully controlled by their control input** (MOSFET, IGBT, GTO, etc.) fall into this category. SCRs and TRIACs are *naturally commutated* 自然换流 – cannot be switched off under gate control – and therefore are not self-commutated.

The symbol for a self-commutated switch is:
The control signal (voltage or current drive) is connected at 'C'. The **diode** shows the direction in which the current can flow.



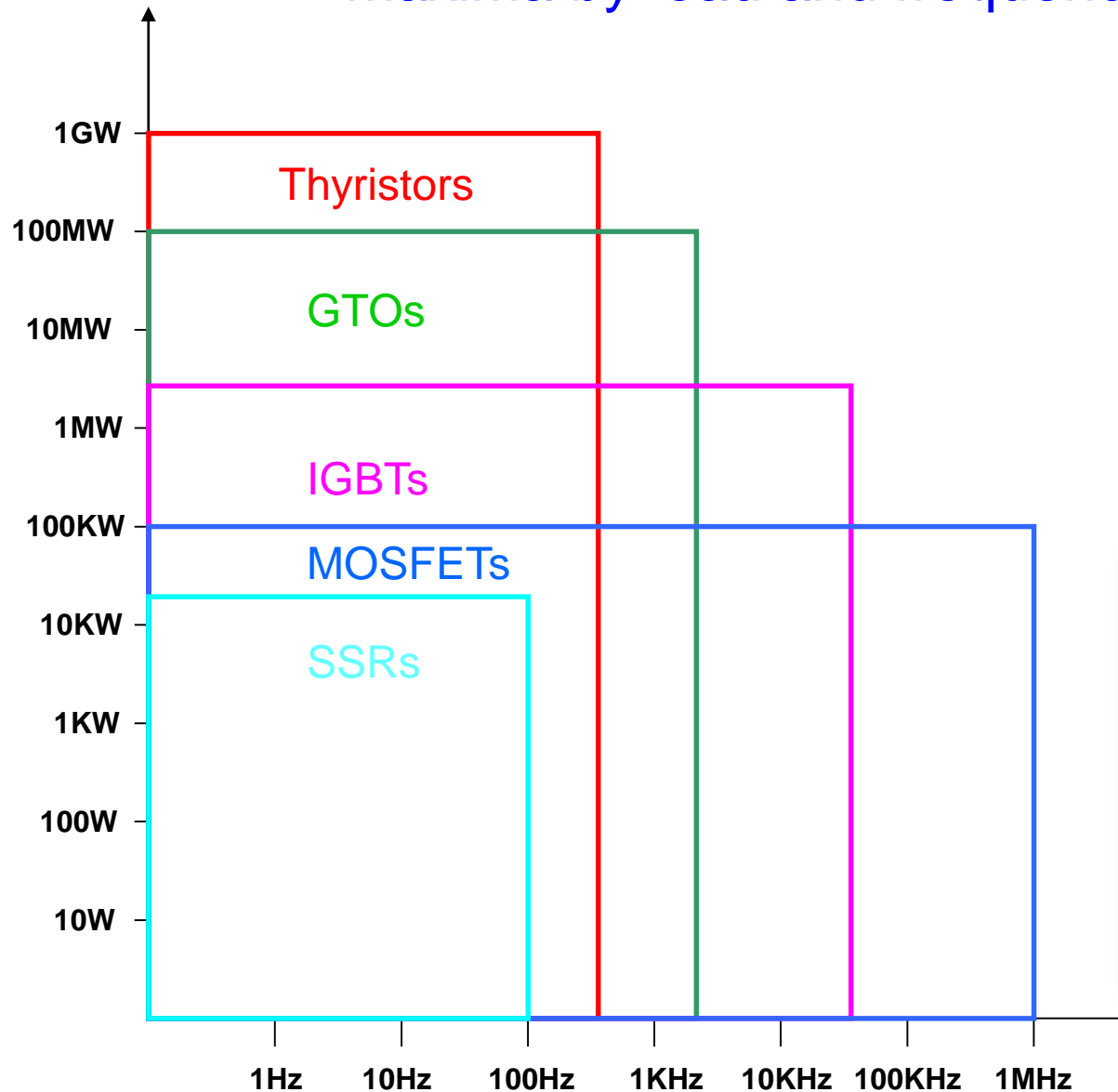
Self-commutated switches are suitable for **use with DC supplies** since they can be turned off independently of the polarity of the source voltage, load voltage or the nature of the load (reactive or resistive).

Three-phase inverters



Electronics power switching

– maxima by load and frequency ratings



IGBT Comparison Table

Device Characteristic	Power Bipolar	Power MOSFET	IGBT
Voltage Rating	High <1kV	High <1kV	Very High >1kV
Current Rating	High <500A	Low <200A	High >500A
Input Drive	Current, h_{FE} 20-200	Voltage, V_{GS} 3-10V	Voltage, V_{GE} 4-8V
Input Impedance	Low	High	High
Output Impedance	Low	Medium	Low
Switching Speed	Slow (μ S)	Fast (nS)	Medium
Cost	Low	Medium	High