Introduction to Modern Power Electronics

Chap.-3 Supplementary Components & Systems

Dr. U. T. Shami

3.1 What Are Supplementary Components & Systems?

A practical power electronic converter is a complex system comprised of several subsystems and numerous components. Many of them are not shown in converter circuit diagrams, which are usually limited to the power circuit and, sometimes, a block diagram of the control system. The supplementary components and systems for modern power electronic converters include:

1. *Drivers* for individual semiconductor power switches, which provide switching signals, interfacing the switches with the control system.
2. *Overcurrent protection schemes*, which safeguard converter switches and sensitive loads from excessive currents.
3. *Snubbers*, which protect switches from transient overvoltages and overcurrents at turn-on and turn-off, and reduce the switching losses.
4. *Filters*, which improve the quality of the power drawn from the source and that supplied to the load. As parts of the power circuit, filters are usually shown in circuit diagrams of converters.
5. *Cooling systems*, which reduce thermal stresses on switches.
6. *Control systems*, which govern the converter operation, including protection tasks.

3.2 Drivers

driver configurations are employed in power electronic converters. A driver, activated by a logic-gate-level signal from the control system, must be able to provide a sufficiently high voltage or current to the controlling electrode (gate or base) to cause immediate turn-on. The on-state of the switch must be safely maintained until turn-off.

Generally, the low-voltage control system and the high-voltage power circuit must be isolated electrically. This can be realized using pulse transformers or optical coupling. The latter is accomplished by placing a light-emitting diode in the vicinity of a light-activated semiconductor device. Alternatively, instead of transferring light signal through free space, a fiber-optic cable can be used. Because of the fundamentally different driving signal requirements, different solutions are used for semicontrolled thyristors (SCRs, triacs, BCTs), current-controlled switches (GTOs, IGCTs, power BJTs), and voltage-controlled hybrid devices (power MOSFETs and IGBTs).

3.2.1 Drivers for SCRs, Triacs, and BCTs

To fire, that is to trigger an SCR, triac, or BCT into conduction, the pulse of the gate current, i_G, must have sufficient magnitude and duration and a possibly short rise time (i.e., high di_G/dt). Isolation is necessary between the control circuitry and the power circuit, at least for switches with ungrounded cathodes. The isolation can be provided by either an optocoupler or a transformer. Both solutions have advantages and disadvantages. An optocoupler requires a power supply and an amplifier on the thyristor side, which are not needed if a transformer is used. However, extra circuitry must be employed to avoid saturation of the transformer core.

3.2.1 Drivers for SCRs, Triacs, and BCTs

A driver for an SCR, based on a pulse transformer, PTR, and transistor amplifier, TRA, is shown in Figure 3.1. Diode D1 and zener diode DZ connected across the primary winding provide a freewheeling path for the primary current at turn-off and prevent saturation of the transformer core. Diode D2 in the gate circuit rectifies the secondary current of the transformer.

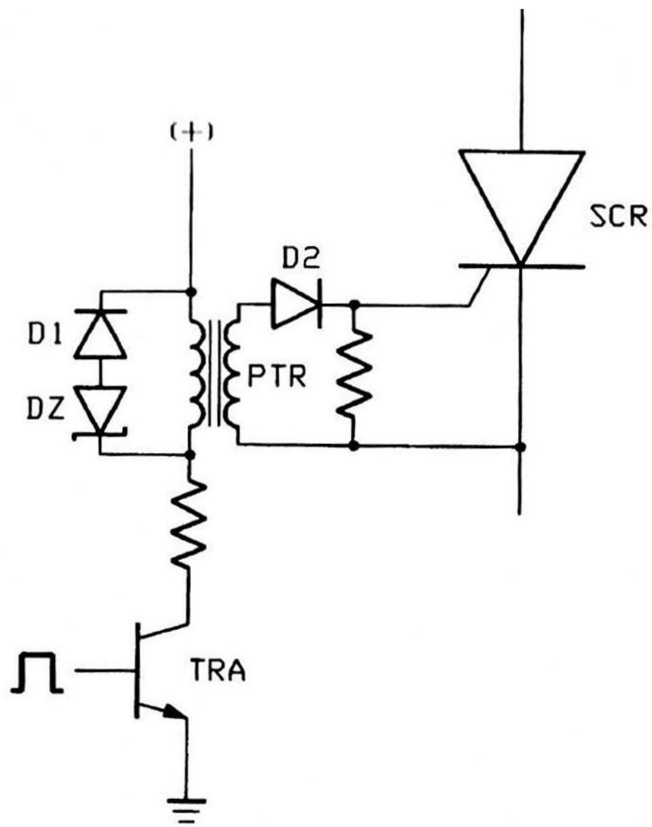


Figure 3.1 Driver for an SCR with transformer isolation.

3.2.1 Drivers for SCRs, Triacs, and BCTs

A simple optically isolated driver for an SCR is shown in Figure 3.2. The optocoupler is comprised of a light-emitting diode (LED) and a small light-activated thyristor (LAT). The energy for the gate signal is obtained directly from the power circuit, as it is the voltage across the SCR that produces the gate current when the LAT is activated by the LED. The LAT must withstand the same voltage as the driven SCR. This is not a serious problem, though, as light-activated thyristors, also used

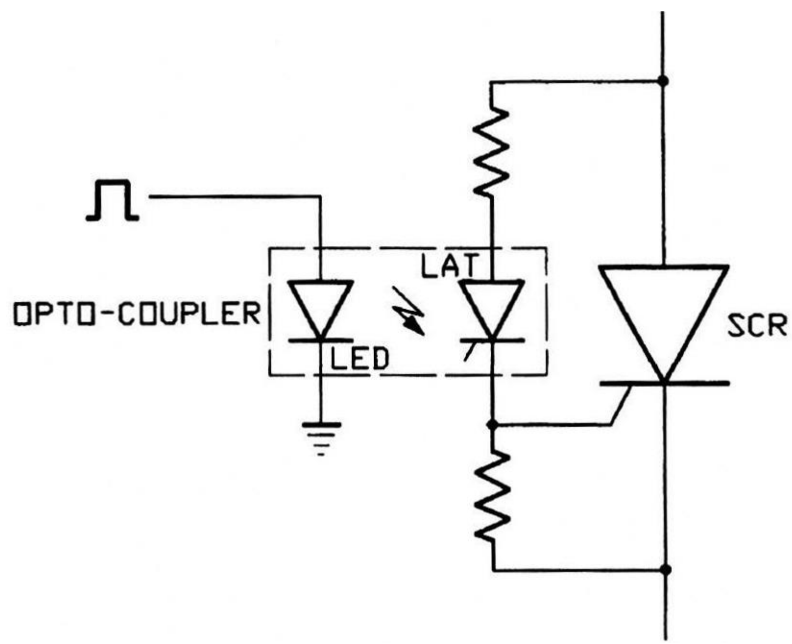


Figure 3.2 Optically isolated driver for an SCR.

3.2.1 Drivers for SCRs, Triacs, and BCTs

Figure 3.3 shows a nonisolated driver for a triac. A transistor amplifier (TRA) provides the gate current for the triac. An optically isolated driver using a light-

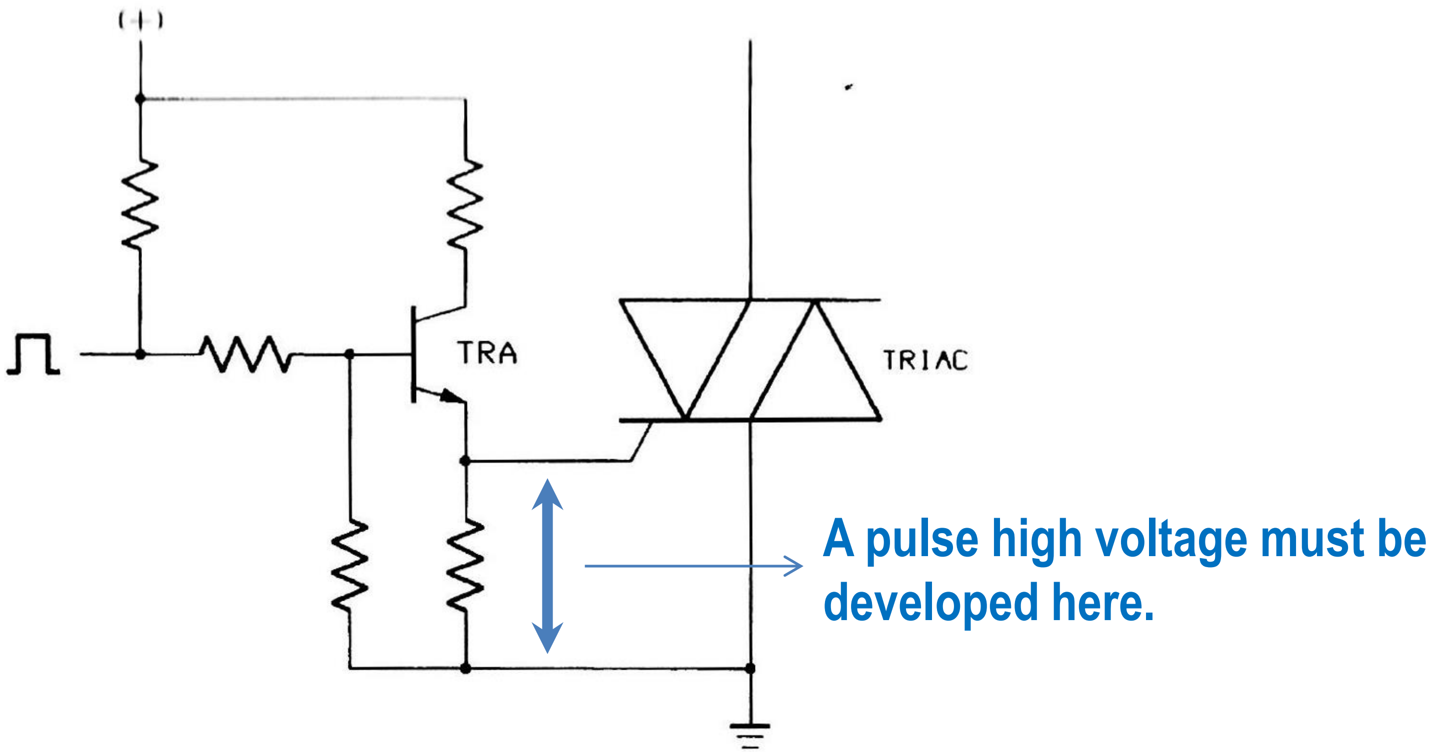


Figure 3.3 Nonisolated driver for a triac.

3.2.1 Drivers for SCRs, Triacs, and BCTs

provides the gate current for the triac. An optically isolated drivèr using a light-activated triac (LATR) is illustrated in Figure 3.4.

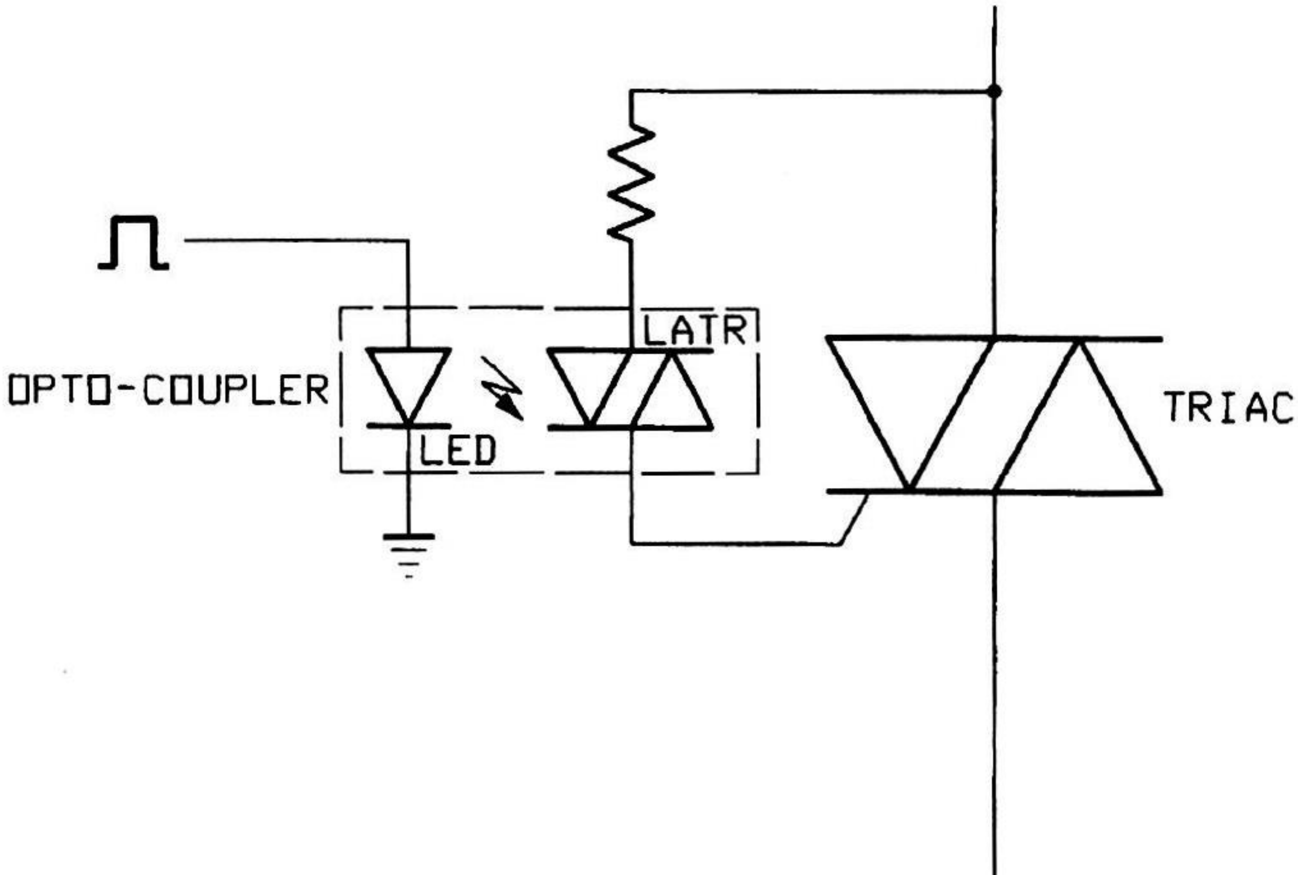


Figure 3.4 Optically isolated driver for a triac.

3.2.2 Drivers for GTOs and IGCTs

Although GTOs and IGCTs are turned on similarly to SCRs, drivers for these switches are more complex than those for the semicontrolled thyristors, because of the very high gate current pulse magnitude required for turn-off. A gate drive circuit is shown in Figure 3.5. To turn the GTO on, the pulse transformer, PTR, transmits high-frequency current pulses generated by alternately switched MOSFETs M1 and M2. The firing current is supplied to the gate through the zener diode, DZ, and inductor, *L*, which controls the rate of change, di_G/dt, of the current. Simultaneously, capacitor C is charged via the four-diode rectifier, RCT.

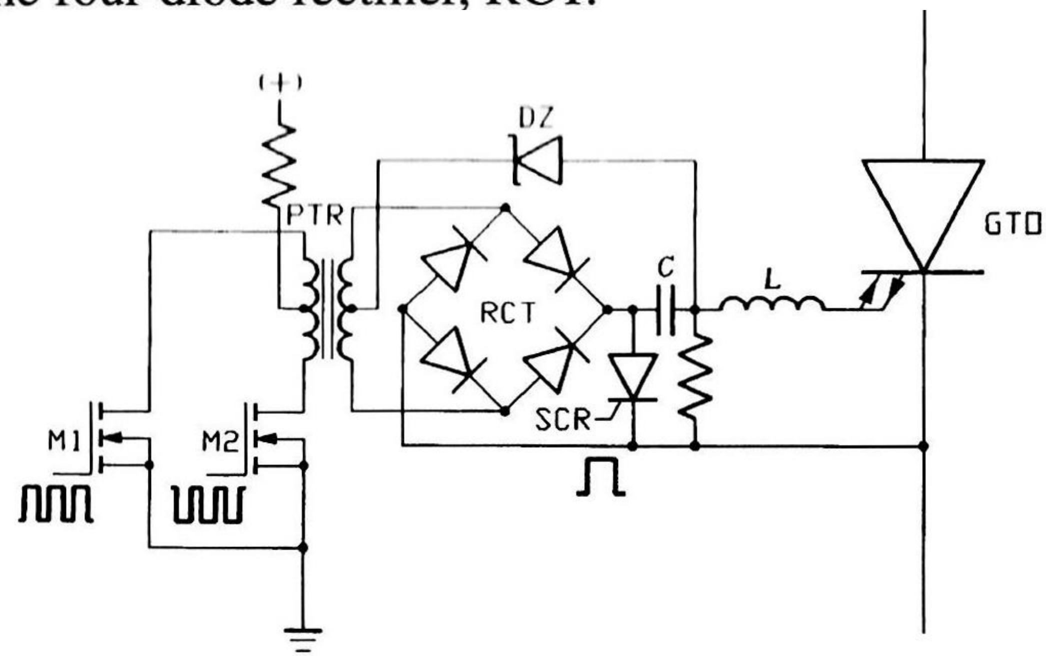


Figure 3.5 Driver for a GTO with transformer isolation.

GTO and IGCT Turn Off Process

A PAUSE OR INTERRUPTION

Cessation of the impulse train indicates that turn-off is to be performed. Turn-off is initiated by the SCR, which causes rapid discharge of the capacitor in the gate–cathode circuit.

Drivers with optical isolation are also known. The GTO side of the driver must have its own power supply to provide the necessary gate current, especially for turn-off. Instead of the SCR in the driver in Figure 3.5, a BJT, a power MOSFET, or a combination can be used to initiate turn-on and turn-off gate pulses.

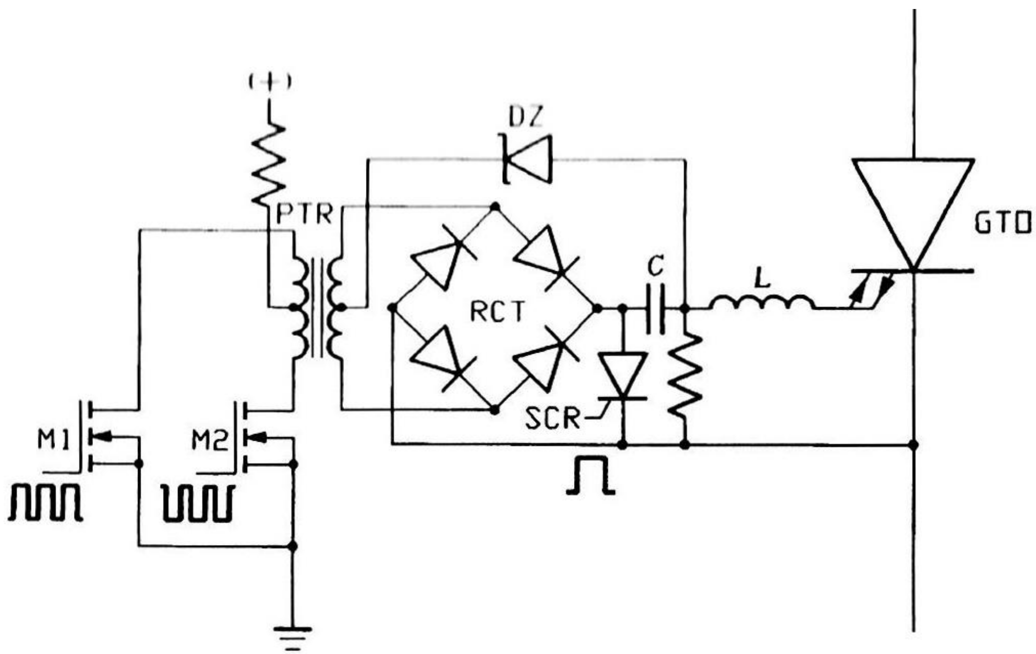


Figure 3.5 Driver for a GTO with transformer isolation.

3.2.3 Drivers for BJTs

Because of the properties of BJTs, base drive circuits must be of the current-source type. A high-quality driver should have the following characteristics:

1. A high current pulse at turn-on, to minimize the turn-on time.
2. An adjustable base current in the on-state, to minimize losses in the base–emitter junction. The initial boost current should be reduced after turn-on.
3. Prevention of hard saturation of the transistor. A saturated BJT has a significantly longer turn-off time than a quasi-saturated BJT.
4. A reverse base current for turn-off, to further minimize the turn-off time.
5. Possibly a low impedance between the base and the emitter in the on-state and a reverse base–emitter voltage in the off-state. These measures increase the collector–emitter voltage blocking capability of the transistor.

3.2.3 Drivers for BJTs

Two simple nonisolated drivers are shown in Figure 3.6. The single-ended driver shown in Figure 3.6a requires only one transistor, TR, but its performance is inferior

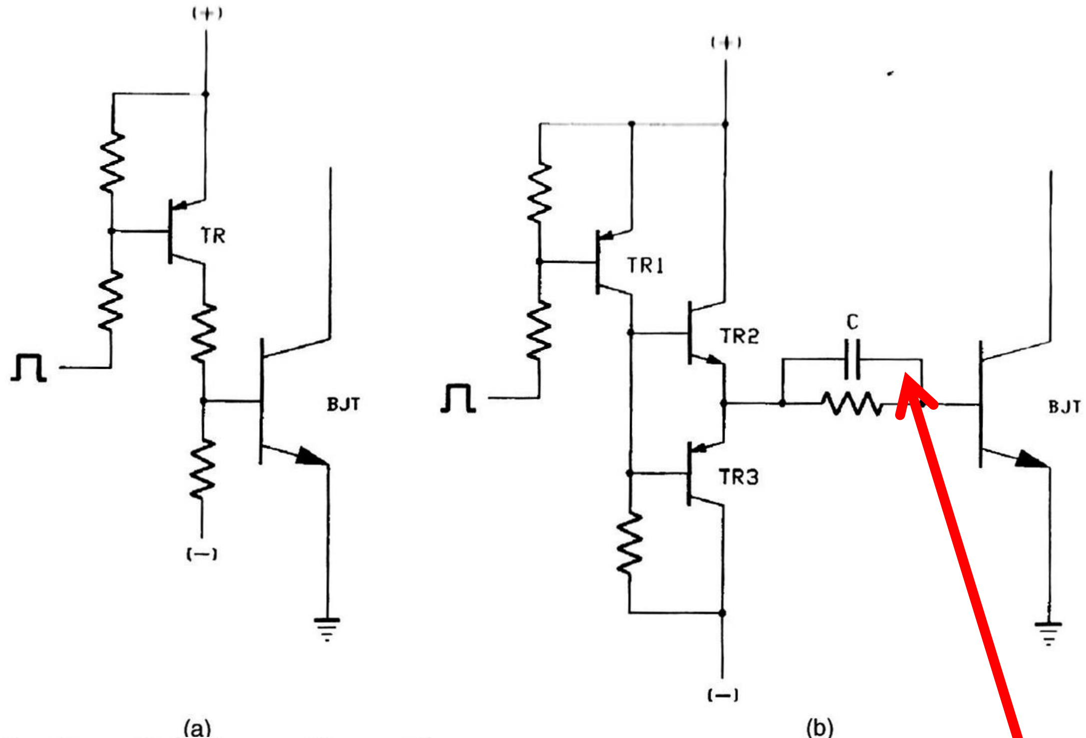


Figure 3.6 Nonisolated drivers for a BJT: (a) single-transistor driver; (b) driver with a class B output stage.

to that of more advanced schemes. Power losses in the driver are reduced in the circuit shown in Figure 3.6b. Input transistor TR1 drives a *class B output stage*, consisting of an npn transistor (TR2) and a pnp transistor (TR3). Capacitor C in both drivers speeds up switching processes, providing a current boost to the base.

Isolated BJT Driver

A commercial single-chip driver with optical isolation

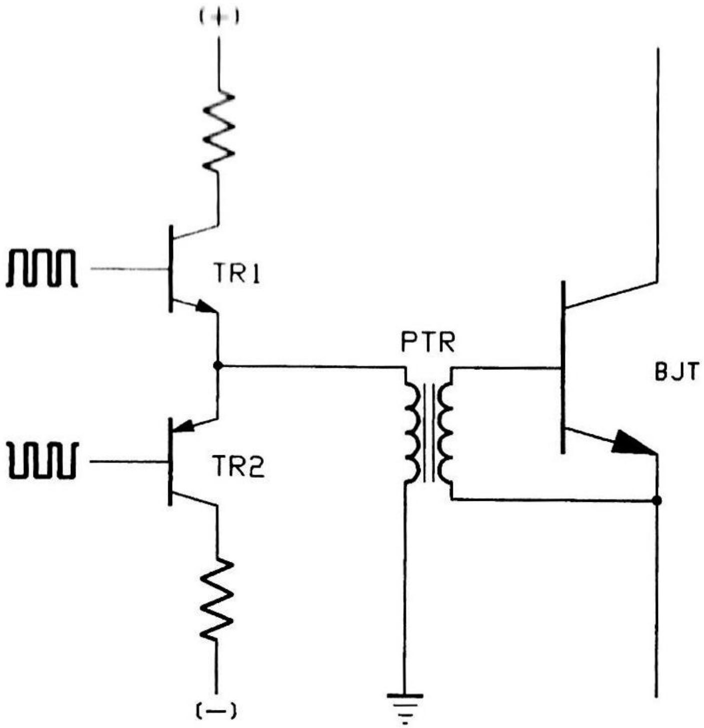
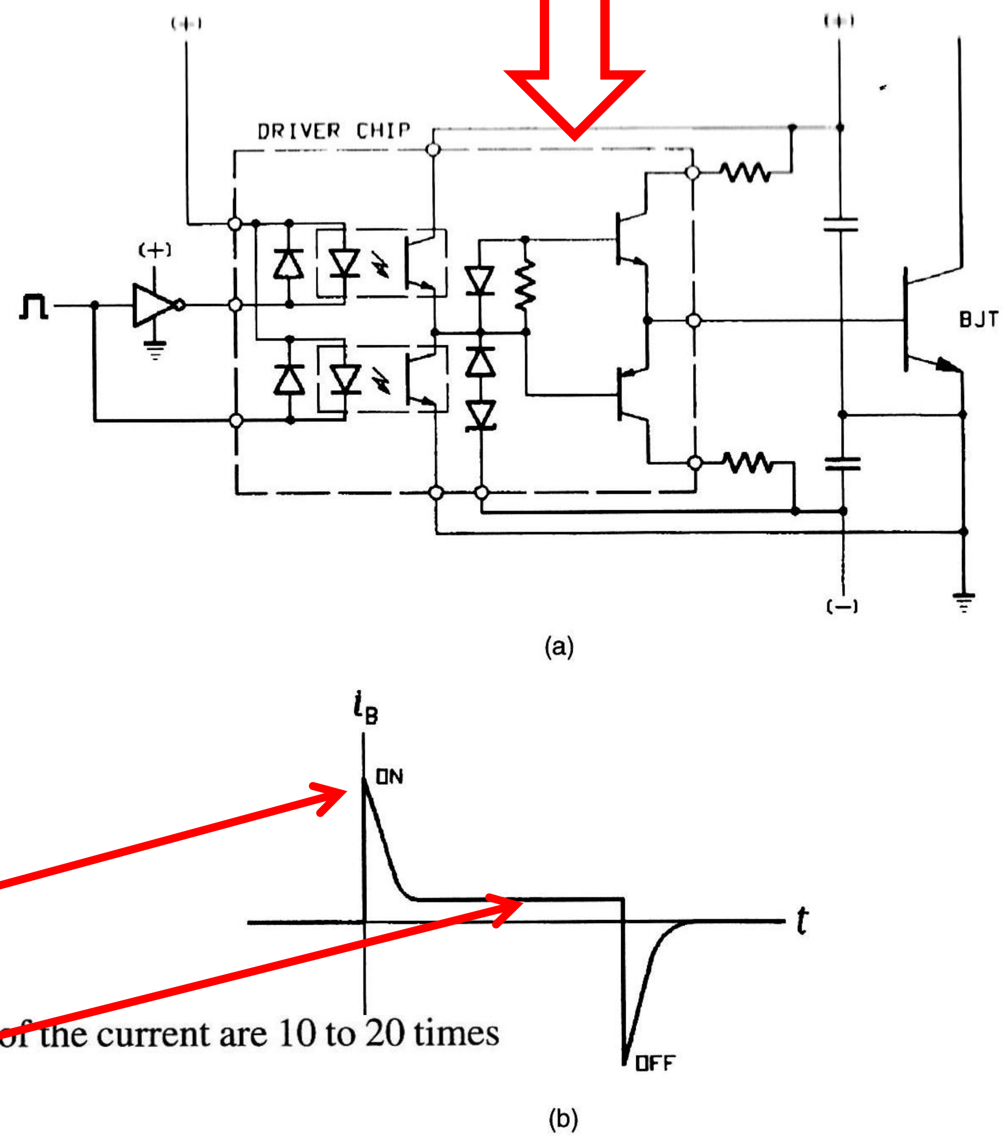


Figure 3.8 Driver for a BJT with transformer isolation.

Technique can be used for High Side.

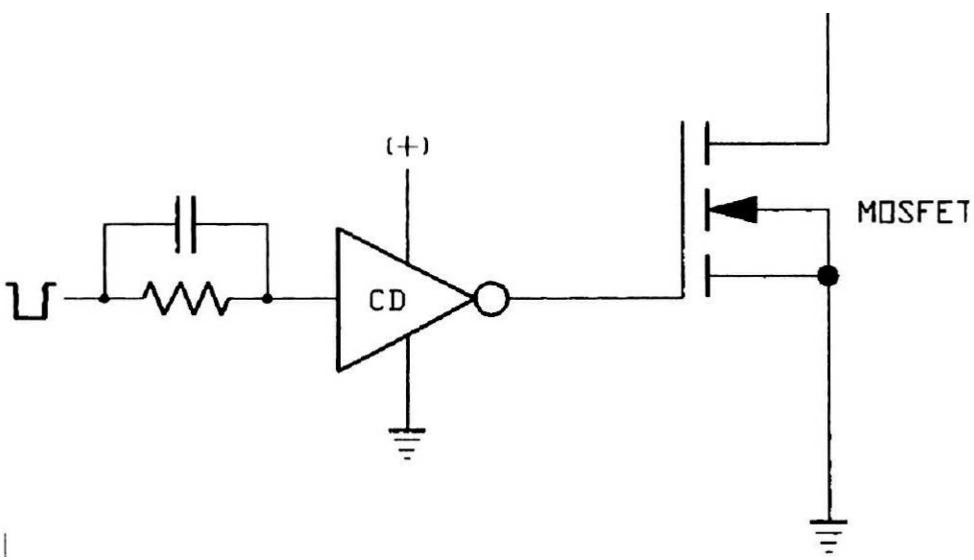


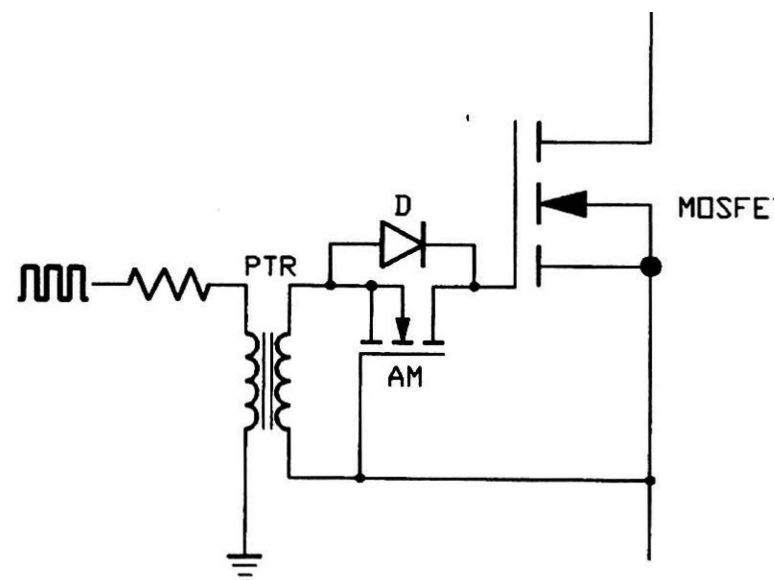
3.9b. In reality, the positive and negative peak values of the current are 10 to 20 times higher than those of sustained on-state current.

Figure 3.9 Driver for a BJT with optical isolation; (a) circuit diagram; (b) waveform of base current.

3.2.4 Drivers for Power MOSFETs and IGBTs

In the steady state, gates of hybrid semiconductor power switches draw almost no current. As such, they can be activated directly from logic gates. However, if high-frequency switching is desired, an electric charge must quickly be transferred to and from the gate capacitance. This requires high pulses of gate current at the beginning of the turn-on and turn-off signals. Standard logic gates by themselves are incapable of supplying (sourcing) or drawing (sinking) such high currents, so that the maximum available switching frequency is seriously limited. Therefore, to fully utilize the





MOSFET Driver

high-speed potentials of hybrid switches, very fast power MOSFETs in particular, provisions must be made in the drivers to source or sink transient current pulses.

For simplicity, all the subsequent drivers are shown in application to power MOSFETs, although they can also be used for IGBTs. A simple gate drive circuit with a high-current TTL clock driver (CD) is shown in Figure 3.10. Figure 3.11 illustrates a power MOSFET driven from a pulse transformer, PTR. The internal parasitic diode, D, in the auxiliary MOSFET, AM, provides the path for the charging current of the main MOSFET's gate capacitance. When the pulse transformer saturates, AM blocks the discharge current from the gate until turn-off, which is initiated by a negative pulse from the transformer that turns AM on. The driver is particularly convenient for switches requiring a floating (ungrounded) gate drive.

A driver for power MOSFETs and IGBTs with optical isolation is shown in Figure 3.12. In addition to turning the driven switch on and off, the driver provides overcurrent protection for the switch. The overcurrent condition is detected by sensing the

Optical Isolation MOSFET Driver

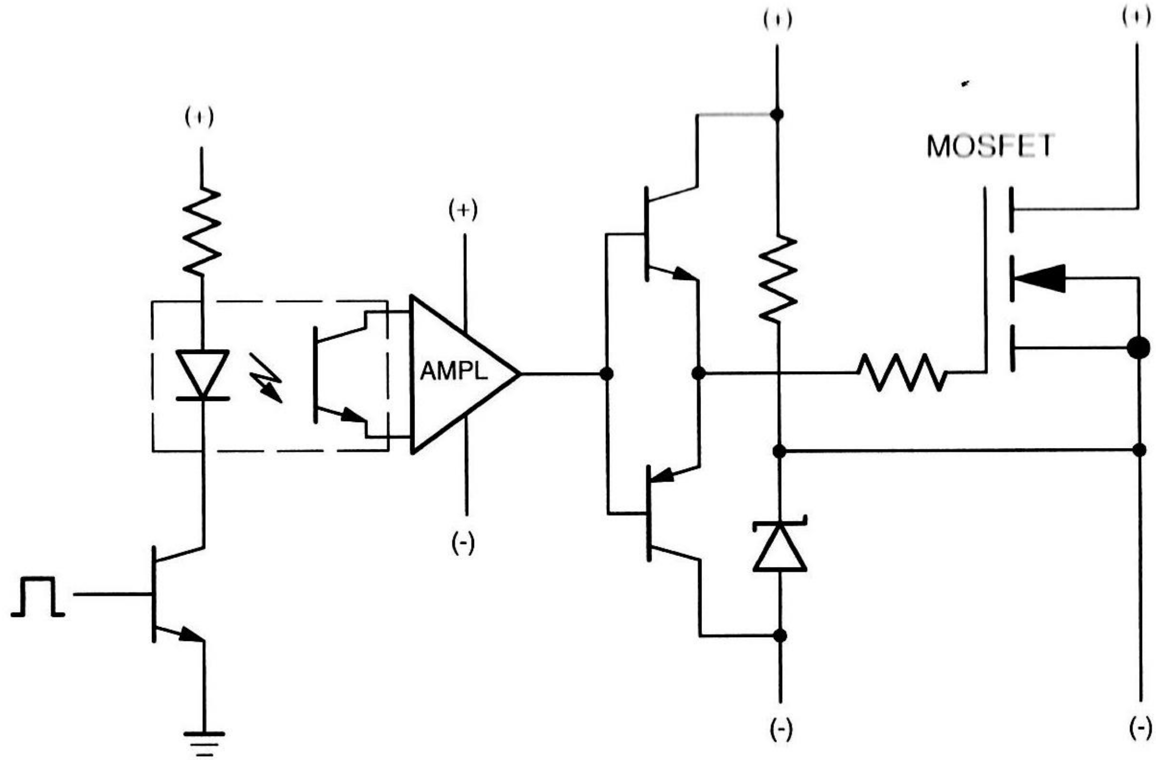


Figure 3.12 Driver for a power MOSFET with optical isolation.

3.3 OVERCURRENT PROTECTION SCHEMES

Semiconductor power switches can easily suffer permanent damage if a short circuit happens somewhere in the converter or the load, or if an overcurrent occurs due to an excessive load, that is, a too-low load impedance and/or counter-EMF. Three basic approaches to overcurrent protection are used in power electronic converters:

1. Fuses
2. SCR "crowbar" arrangement
3. Turning the switches off when an overcurrent is detected.

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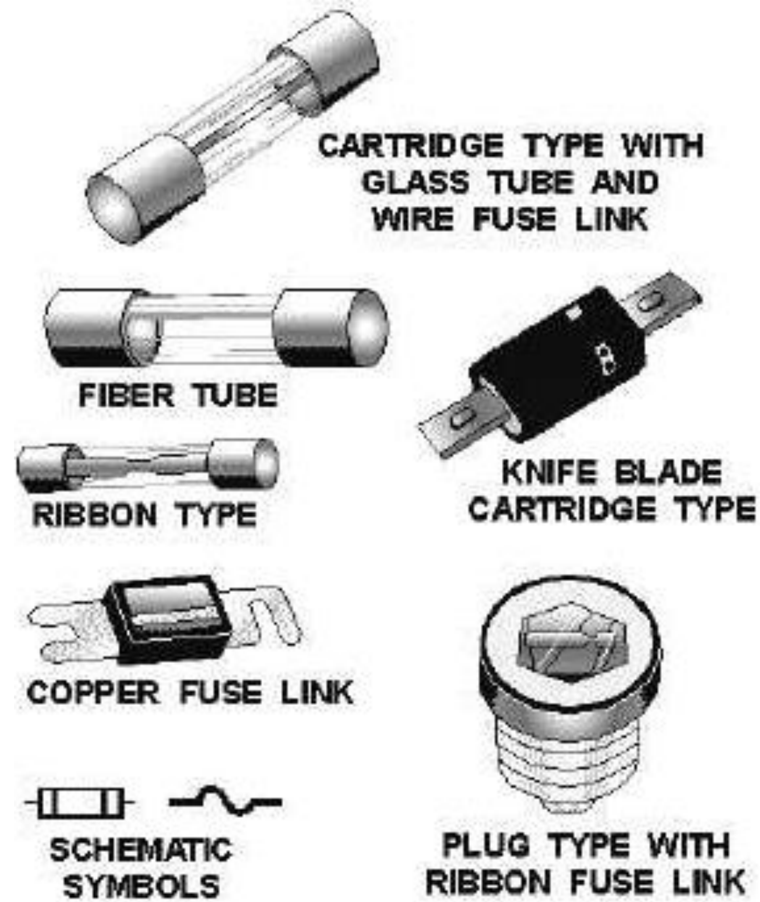
3.3 OVERCURRENT PROTECTION SCHEMES

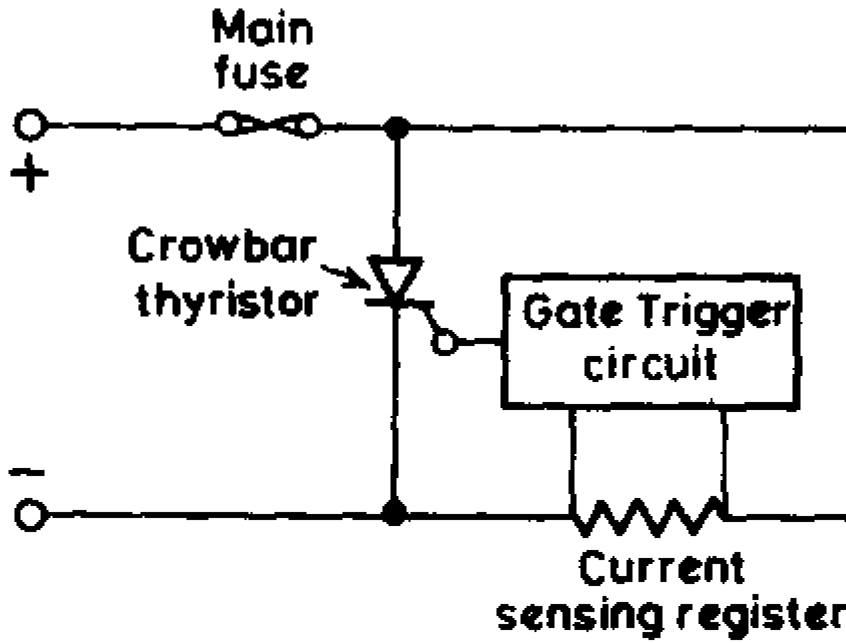
1. Fuses



brackets. The I^2t parameter of a fuse must be less than that of the device being protected but not so low as to cause a breakdown under normal operating conditions. A properly selected fuse should melt within a half-cycle of the 60-Hz (or 50-Hz) voltage.

3.3 OVERCURRENT PROTECTION SCHEMES

2. SCR "crowbar" arrangement



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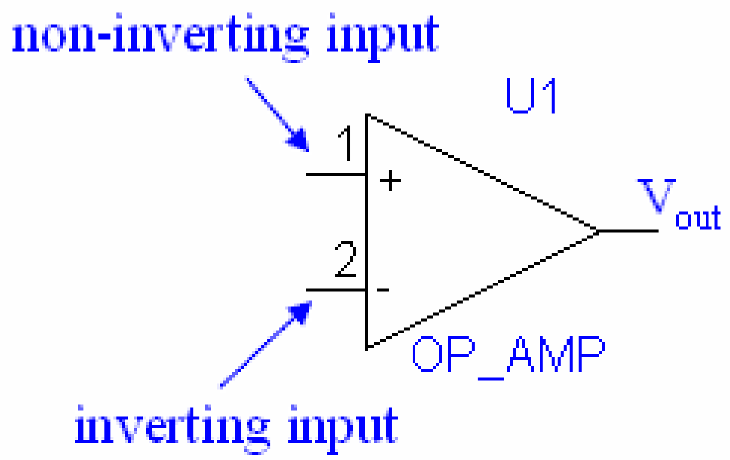
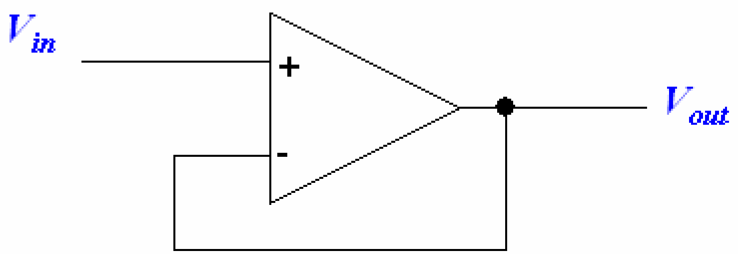
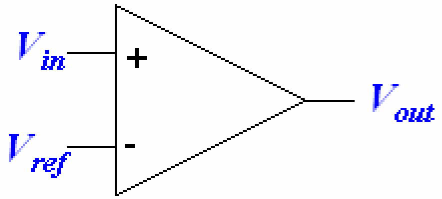


Voltage Follower

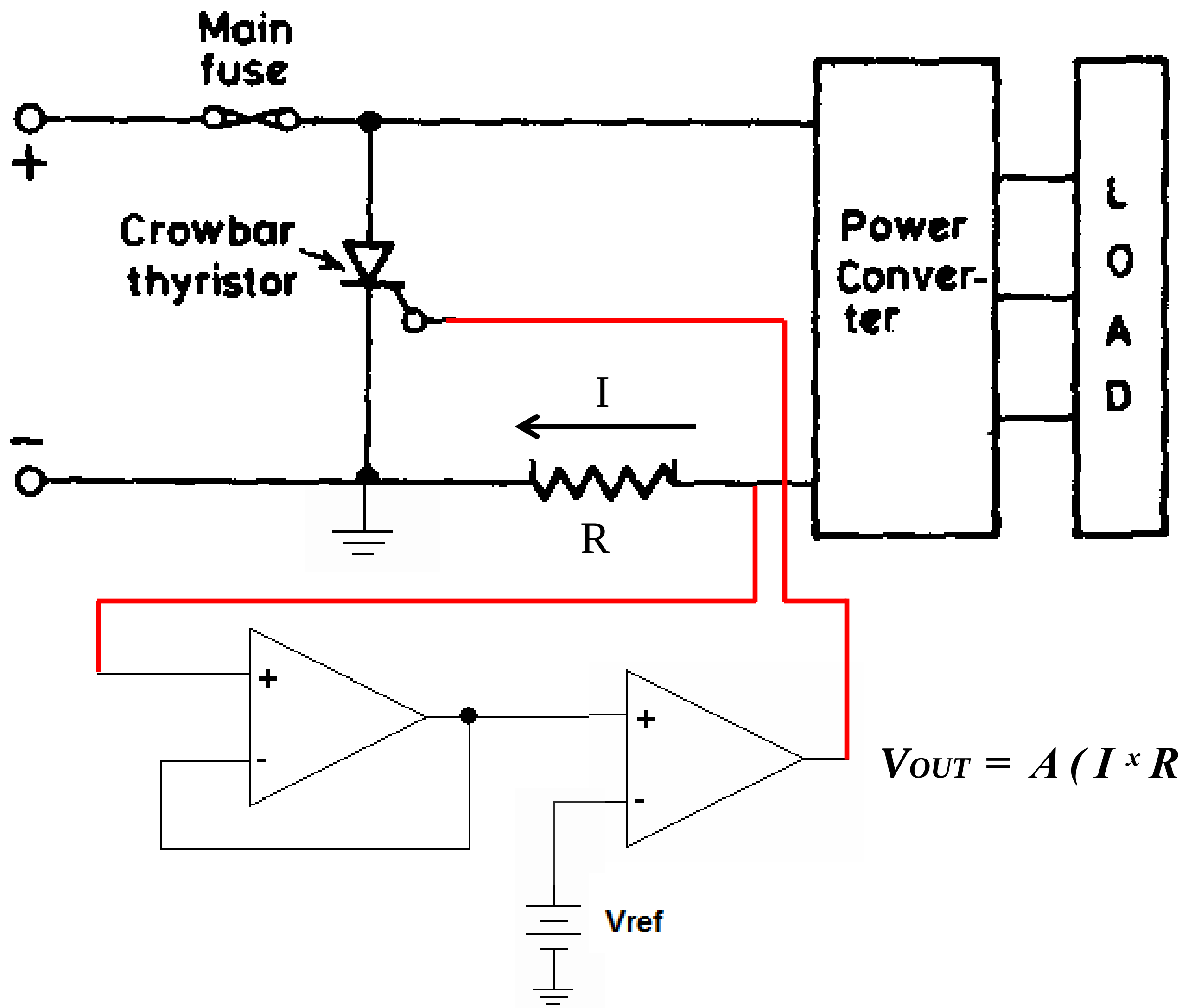


Comparator



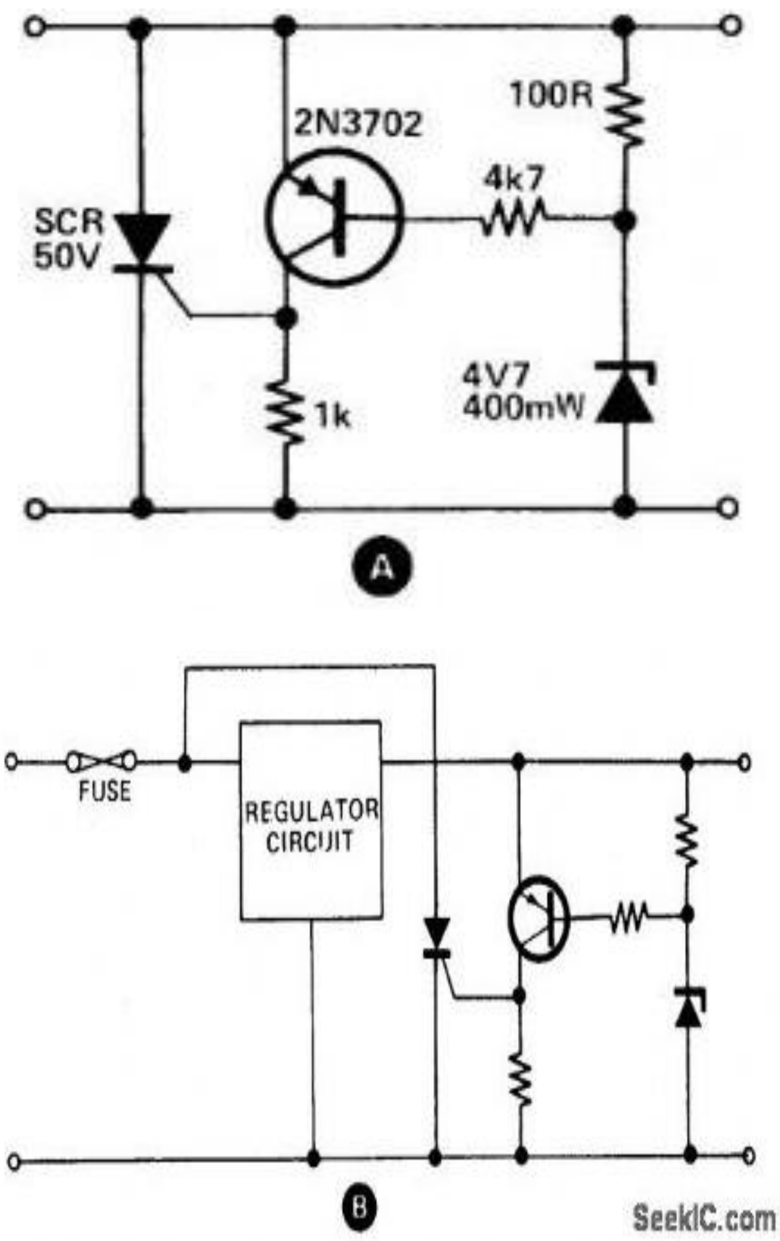
$$V_{out} = A(V^{+} - V^{-})$$

3.3 OVERCURRENT PROTECTION SCHEMES



$$V_{OUT} = A\,(I \times R - V_{REF})$$

Crow-Bar for Over-Voltage Protection



These circuits provide overvoltage protection in case of voltage regulator failure or application of an external voltage. Intended to be used with a supply offering some form of short circuit protection, either foldback, current limiting, or a simple fuse. The most likely application is a 5 V logic supply, since TTL is easily damaged by excess voltage. The values chosen in A are for a 5 V supply, although any supply up to about 25 V can be protected by simply choosing the appropriate zener diode. When the supply voltage exceeds the zener voltage +0.7 V, the transistor turns on and fires the thyristor. This shorts out the supply, and prevents the voltage rising any further. In the case of a supply with only fuse protection, it is better to connect the thyristor the regulator circuit when the crowbar operates. The thyristor should have a current rating about twice the expected short circuit current and a maximum voltage greater than the supply voltage. The circuit can be reset by either switching off the supply, or by breaking the thyristor circuit with a switch.

3.3 OVERCURRENT PROTECTION SCHEMES

3. Turning the switches off when an overcurrent is detected.

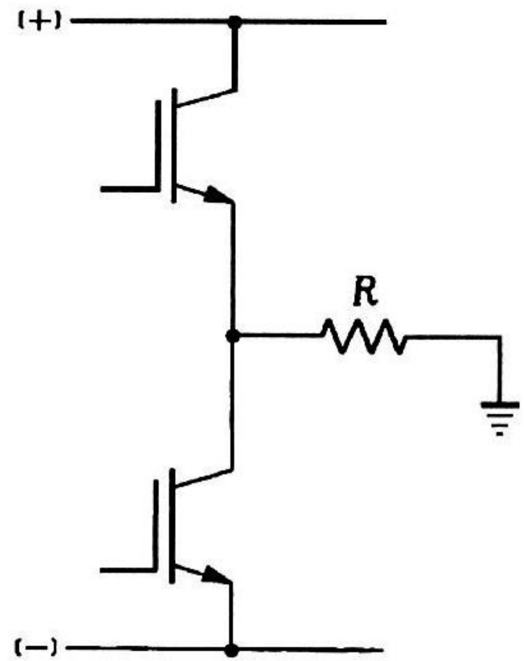


Figure 3.14 Totem-pole arrangement of two switches in a leg of a bridge topology.

3.3 OVERCURRENT PROTECTION SCHEMES

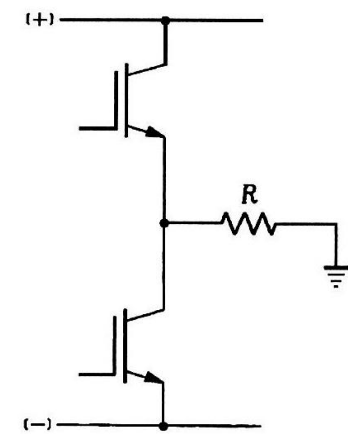
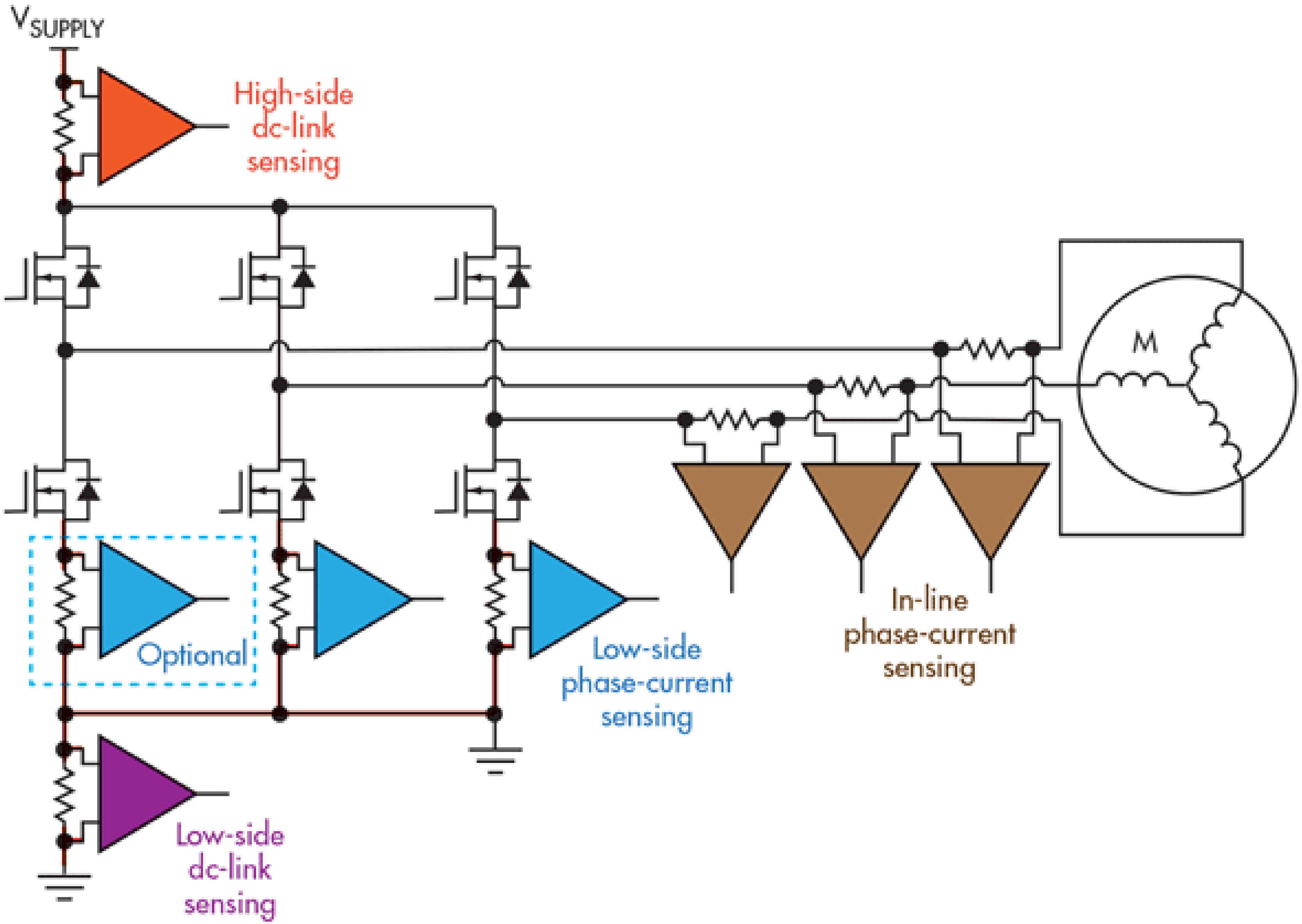


Figure 3.14 Totem-pole arrangement of two switches in a leg of a bridge topology.

A special case of a potentially dangerous short circuit, called a *shot-through*, is specific for the *bridge topology*, typical for many power converters. In individual legs (branches) of a bridge circuit, two (or more) switches are connected in series in a *totem-pole* arrangement, illustrated in Figure 3.14 by two IGBTs. Normally, their states are mutually exclusive; that is, when one switch is on, the other is off, and the current always flows through the load, *R*. However, if for any reason (e.g., due to driver failure or incorrect timing of switching signals) one switch is turned on before the other has turned off, a short circuit occurs. To reduce the probability of an overshoot, the beginning of a turn-on signal for one switch is delayed with respect to the end of a turn-off signal for the other switch by *dead time*.

3.3 OVERCURRENT PROTECTION SCHEMES



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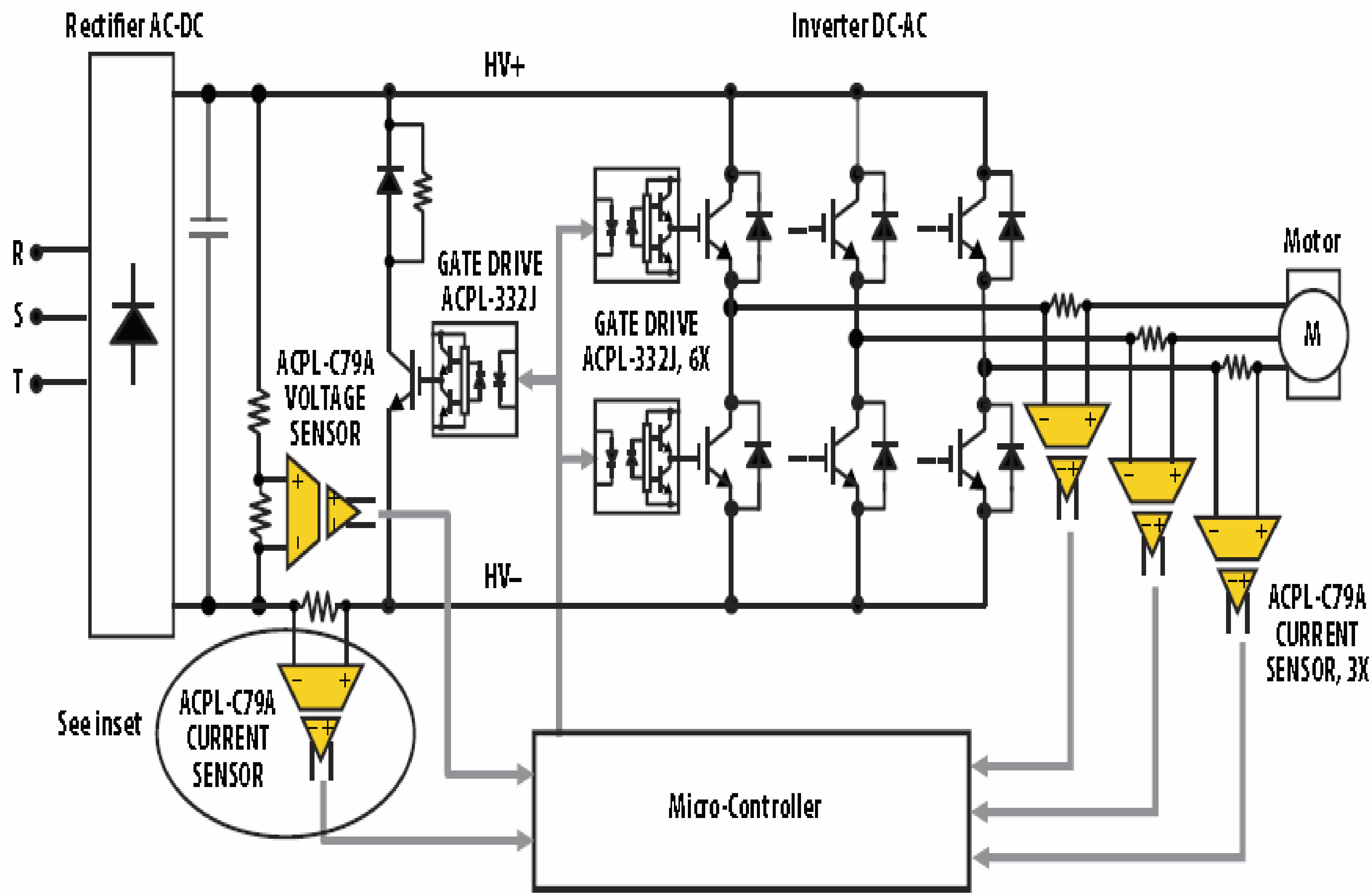
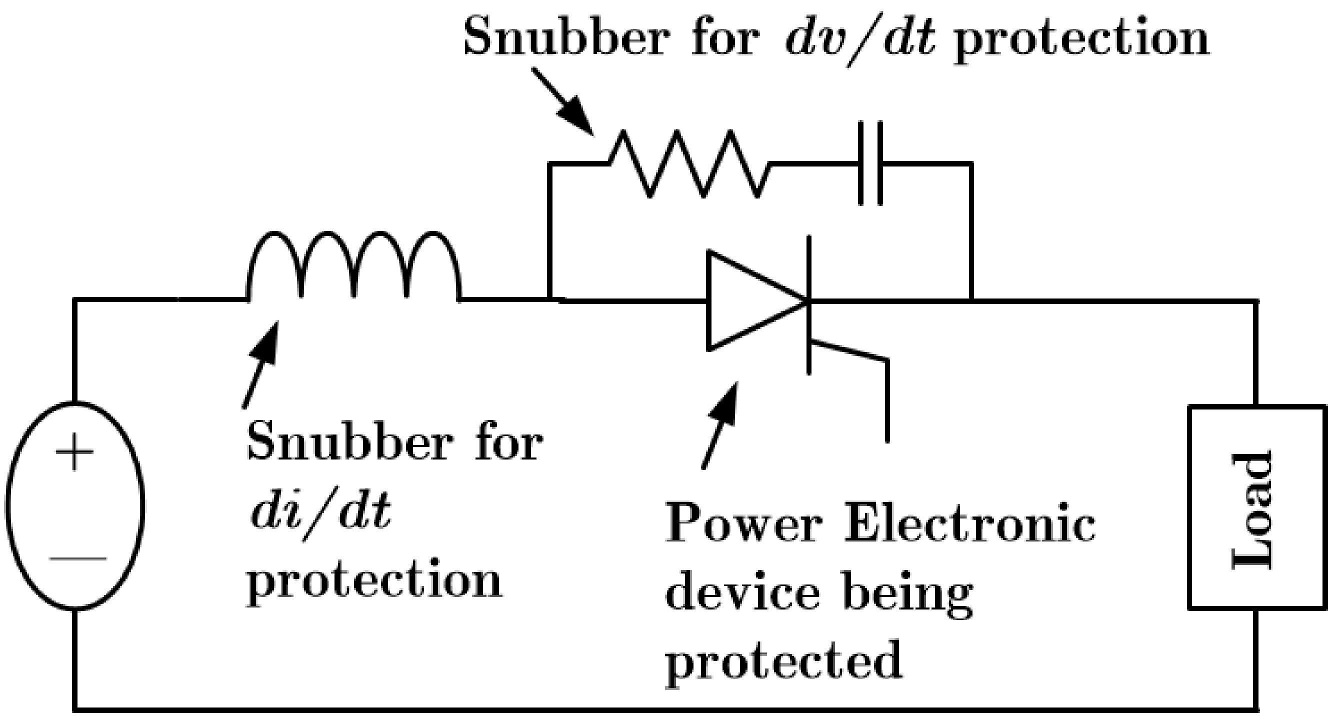


3.4 SNUBBERS

Snubber circuits are needed to limit the rate of change in voltage or current *(di/dt or dv/dt)* and over—voltage during turn-on and turn-off. These are placed across the semiconductor devices for protection as well as to improve the performance.



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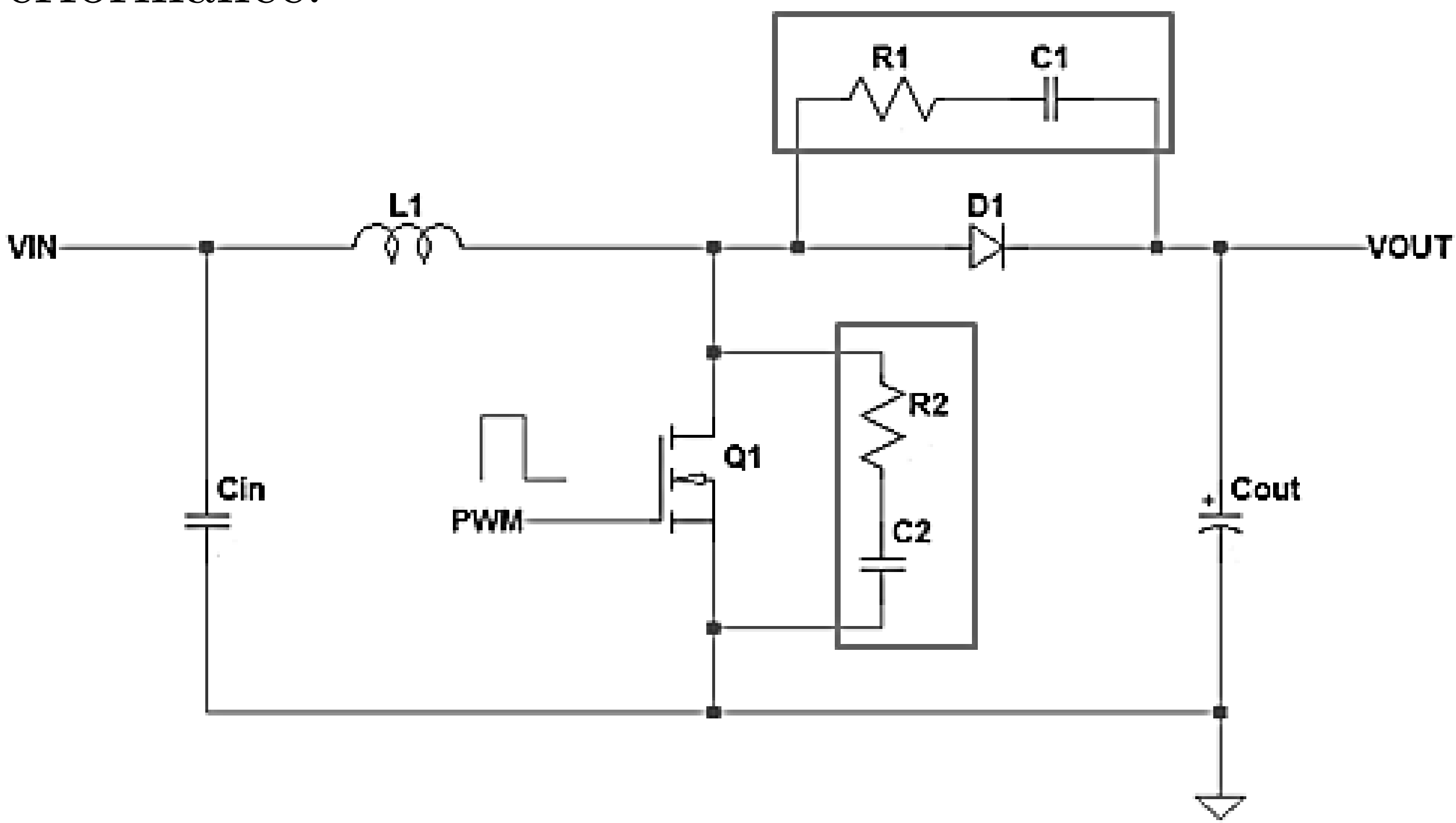
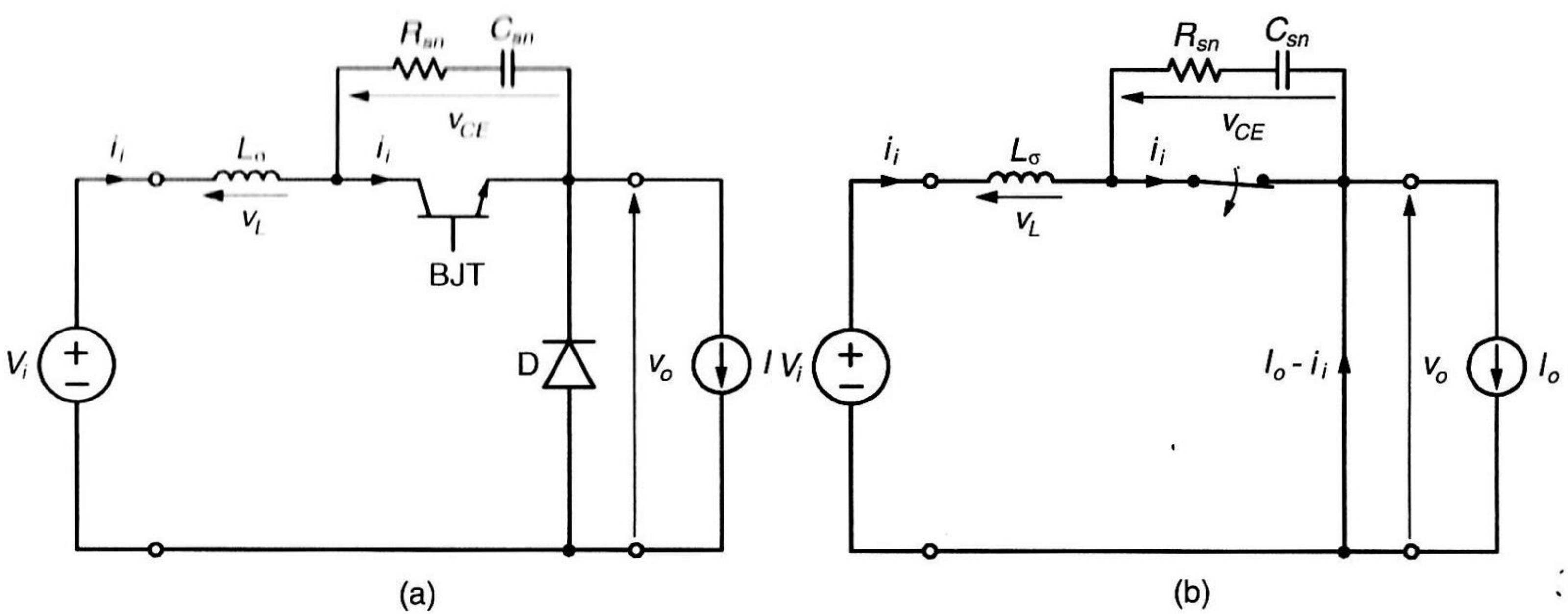


3.4 SNUBBERS

To illustrate the need for snubbers, a practical example is considered. A simple BJT-based chopper is shown in Figure 3.15a. The load inductance is assumed to be so high that the output current is practically constant and equal to I_o. Consequently, the load can be modeled by a current source. A stray inductance, L_σ, of the power circuit of the chopper is lumped between the source of input voltage, V_i, and the transistor. The snubber circuit, composed of resistance R_{sn} and capacitance C_{sn}, is connected in parallel with the transistor.



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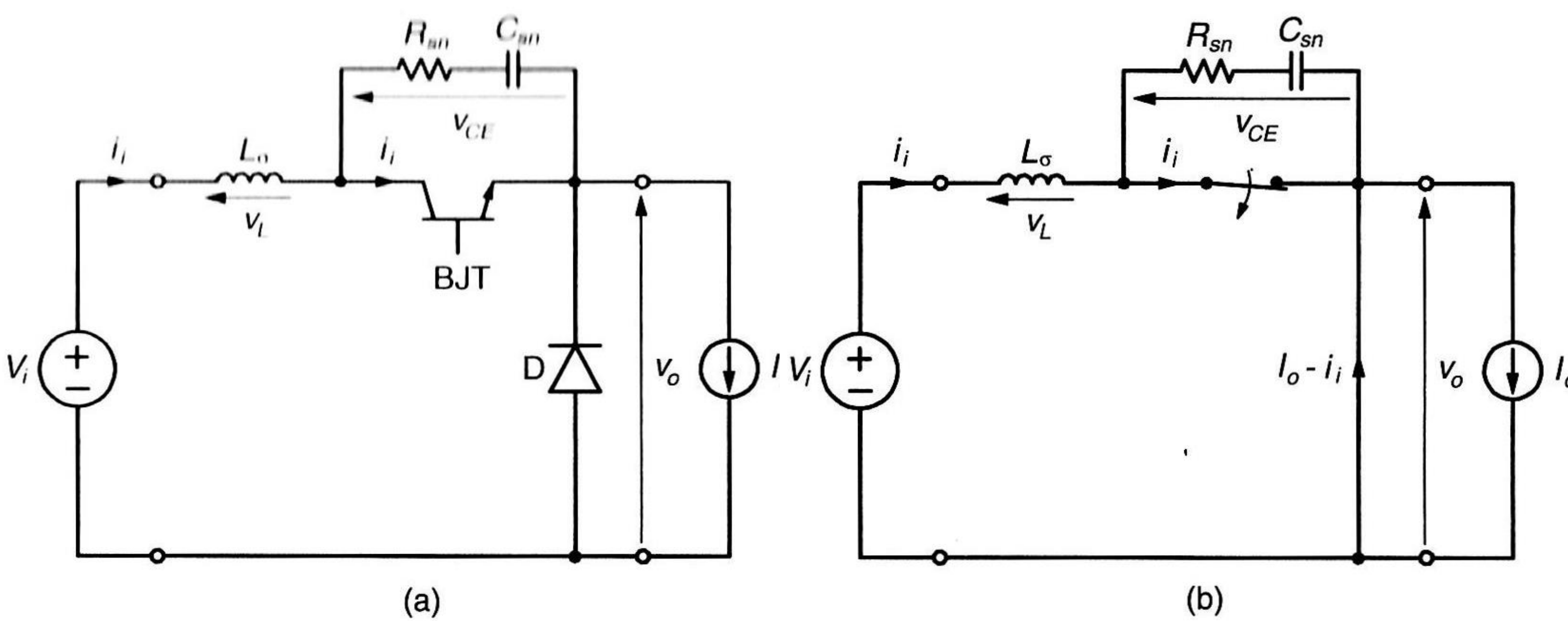

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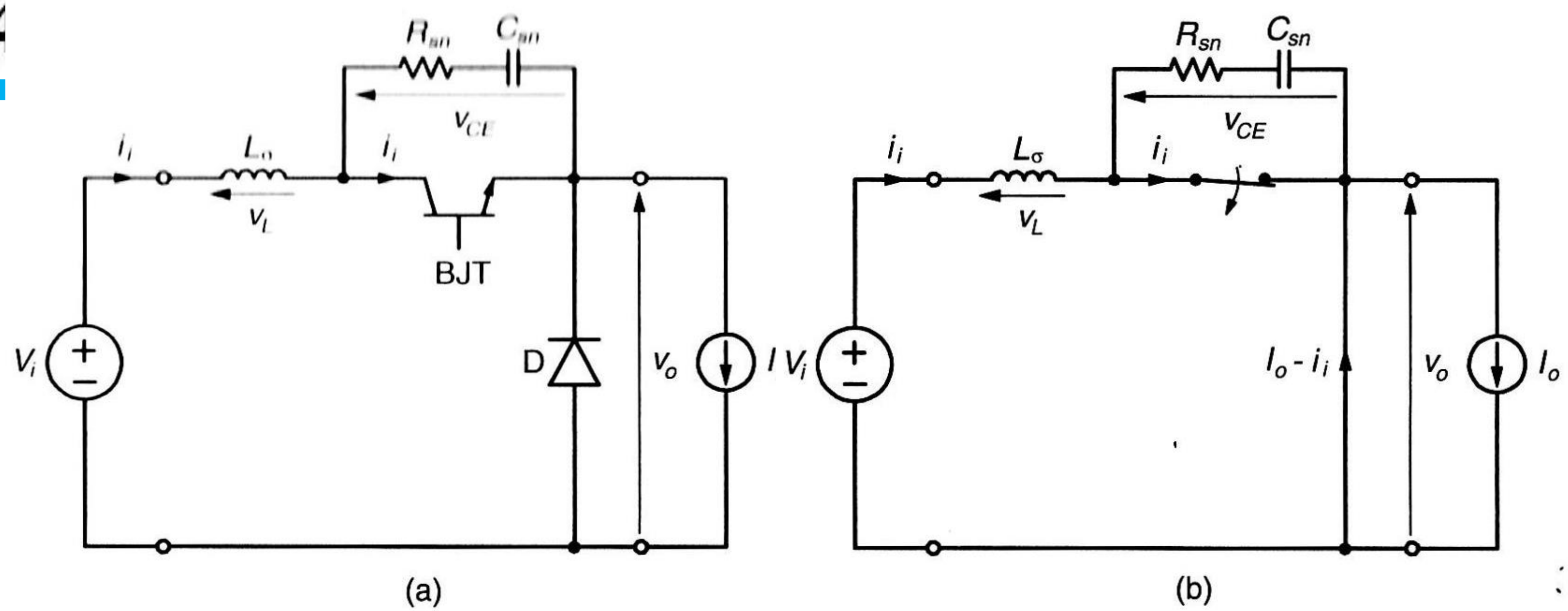


R_sn
C_sn
v_CE
i_i
L_σ
v_L
BJT
V_i
D
v_o
I
(a)
R_sn
C_sn
v_CE
i_i
L_σ
v_L
V_i
I_o - i_i
v_o
I_o
(b)

The collector–emitter voltage, v_{CE}, across the BJT is given by

$$v_{\mathrm{CE}} = V_i - v_{\mathrm{L}} - v_o \tag{3.1}$$

transistor in the on-state, $v_{\mathrm{CE}} \approx 0$. At $t = 0$, the BJT is turned off so that its collector current, i_{C}, decreases linearly from the initial value of I_o, reaching zero at $t = t_0$. As a result, a transient voltage appears across the stray inductance. The voltage waveform has the shape of a pulse with duration (width) t_0 and peak value, $V_{\mathrm{L},p}$, of

$$V_{\mathrm{L},p} = L_\sigma \frac{di_c}{dt} = -L_\sigma \frac{I_o}{t_0}. \tag{3.2}$$

In the meantime, the freewheeling diode, D, has taken over conduction of the output current, thus, $v_o = 0$ and the peak value, $V_{\mathrm{CE},p}$ of the collector–emitter voltage is

$$V_{\mathrm{CE},p} = V_i - V_{\mathrm{L},p} = V_i + L_\sigma \frac{I_o}{t_0}. \tag{3.3}$$

Fast turn-offs are desirable from the point of view of reduction of switching losses, but Eq. (3.3) shows that when t_0 approaches zero, the voltage across the BJT approaches infinity. Even with a finite but short t_0, the voltage can easily be excessive, damaging the transistor.

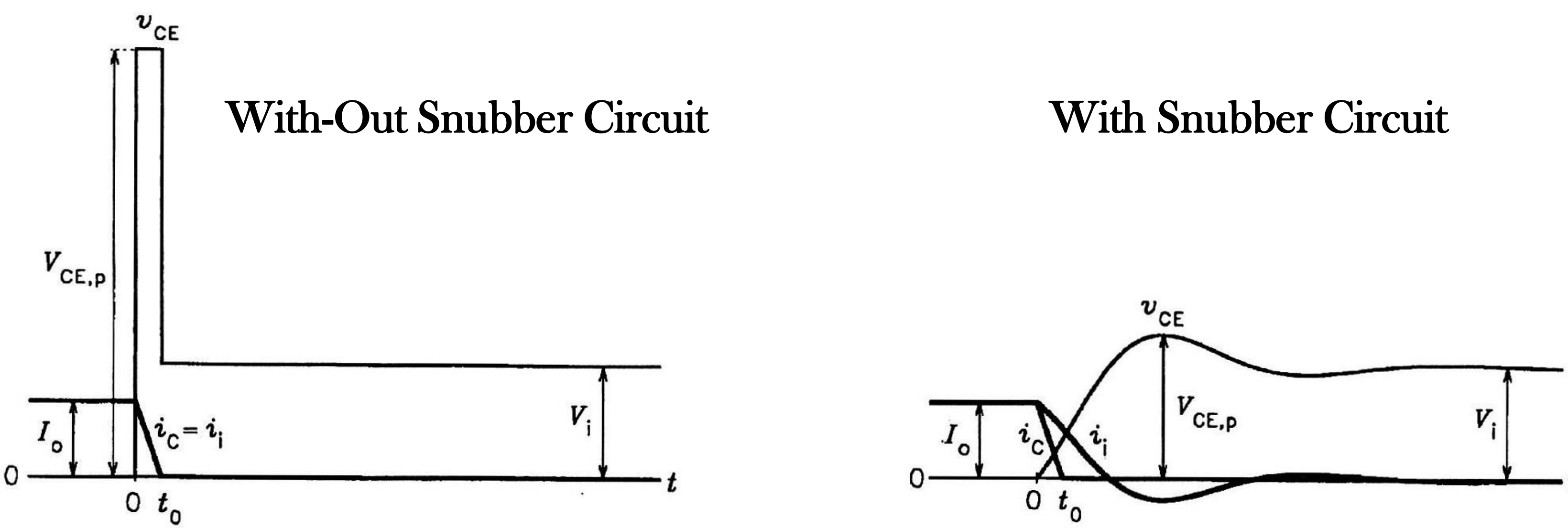
3.4 SNUBBERS

With the snubber in place, the equivalent circuit of the chopper at turn-off is as shown in Figure 3.15b. When switch S representing the BJT opens, a series RLC circuit is created. As known from the theory of such circuits, if $R_{sn} < 2\sqrt{L_\sigma / C_{sn}}$, the current, i_i, in the circuit is given by

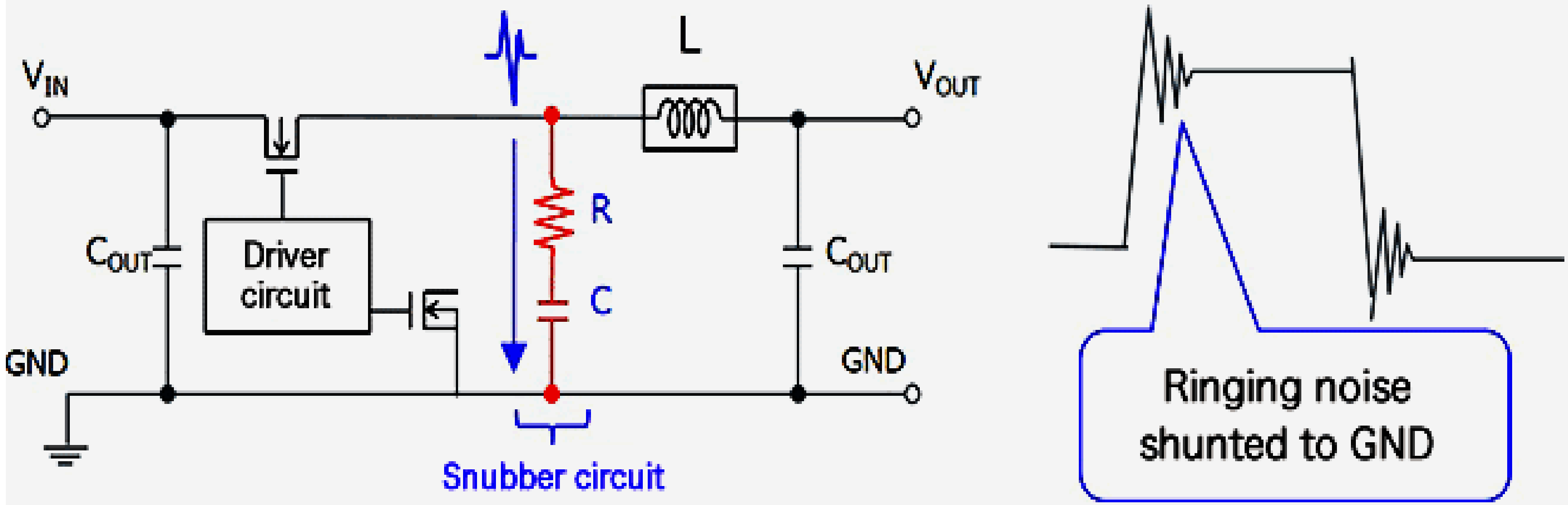
$$i_i = I_p e^{-(R_{sn}/L_\sigma)t} \cos(\omega_d t + \varphi) \tag{3.4}$$

$$v_{CE} = V_i \left[1 - e^{-(R_{sn}/L_\sigma)t} \cos \omega_d t\right]. \tag{3.8}$$

With a properly designed and tuned snubber, the collector–emitter voltage displays only a small overshoot. Voltage and current waveforms for the both cases considered are shown in Figure 3.16. By combining the $i_C(t)$ and $v_{CE}(t)$ waveforms into the



3.4 SNUBBERS

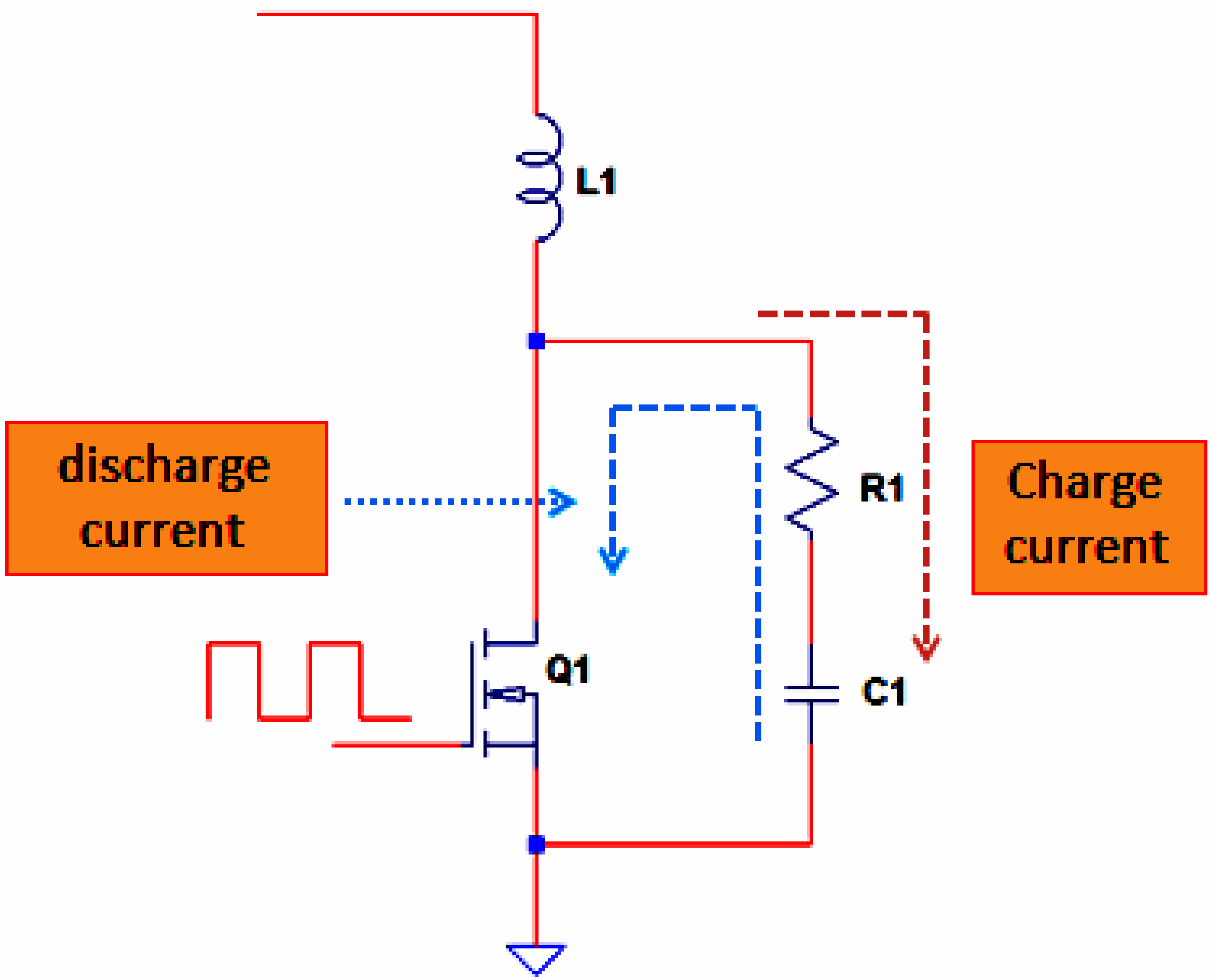


*Effect of snubber circuit depends on trade-off with losses

*High frequencies shunted to GND by C and R

*Attention must be paid to allowable losses of resistor

3.4 SNUBBERS



3.4.2 Snubbers for GTOs and IGCTs

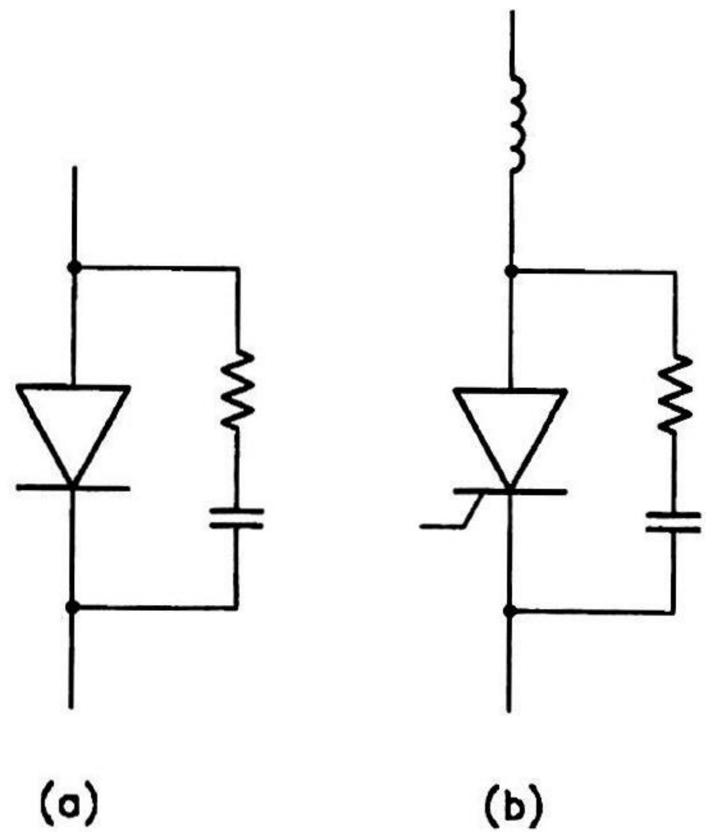


Figure 3.18 Snubbers for (a) a power diode; (b) an SCR.

3.4.3 Snubbers for Transistors

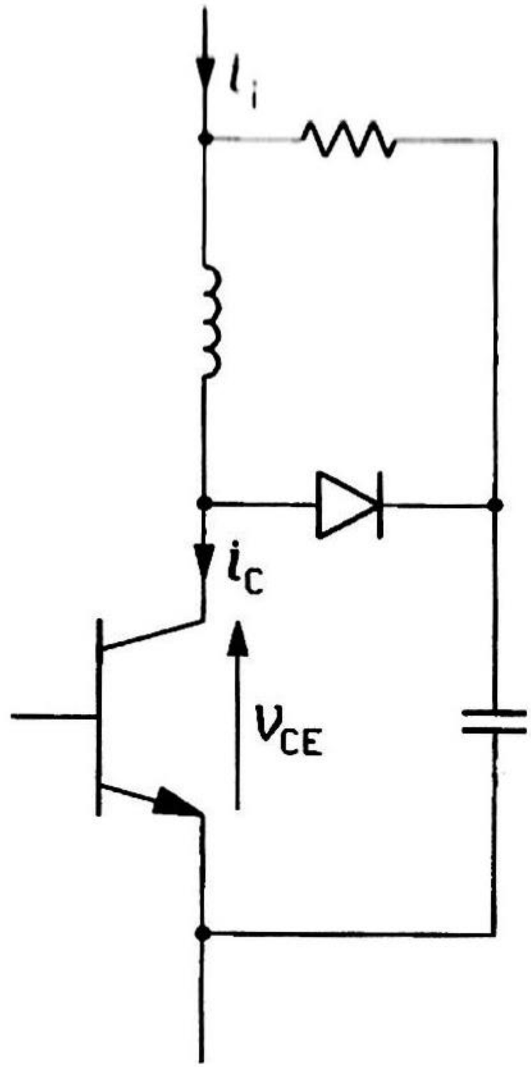
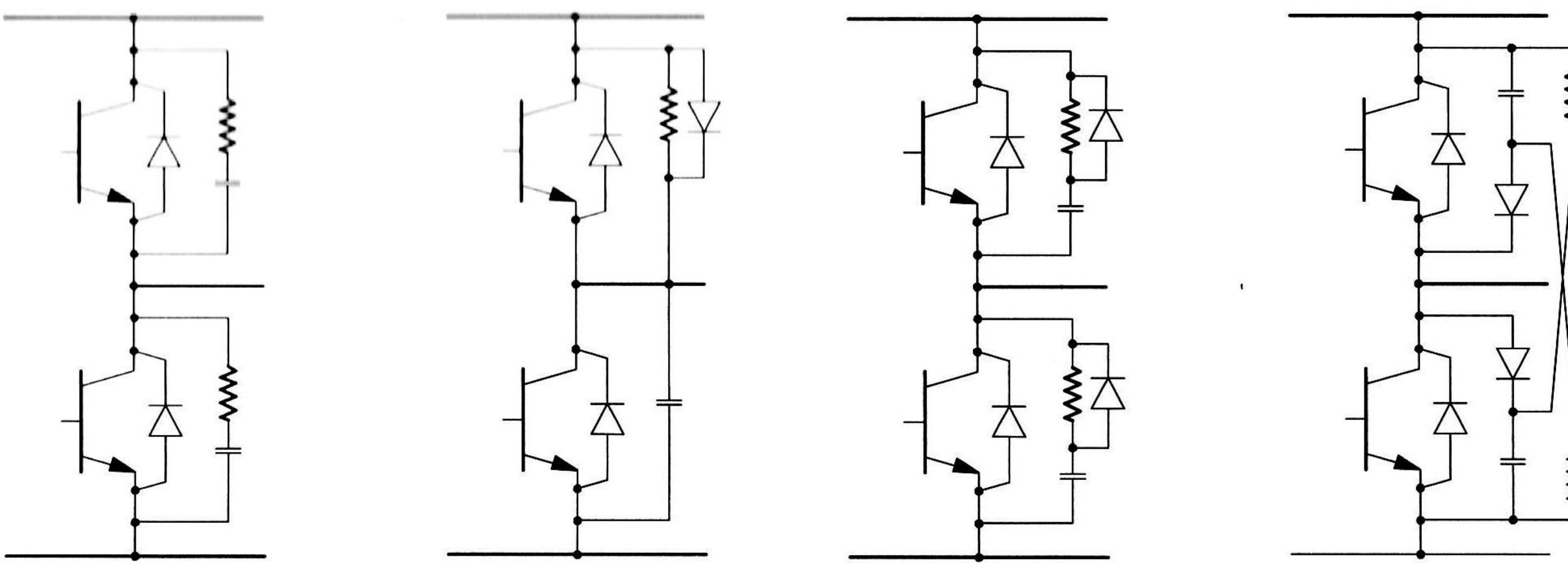


Figure 3.20 Combined on-and-off snubber for a transistor.

3.4.4 Energy Recovery from Snubbers

The snubbers we have discussed modify the switching trajectory and reduce switching losses. However, the energy stored temporarily in the inductive and capacitive components is dissipated in resistors, that is, lost irretrievably. In high-frequency and high-power converters, the amount of energy lost in snubbers is substantial, straining cooling systems and reducing the efficiency of the converters. Therefore, measures have been developed to recover energy from snubbers and direct it to the load or return it to the supply source.



(a) (b) (c) (d)

Baker's Clamp

To increase the turn-off speed, an anti-saturation circuit shown in Figure 3.7 called a *Baker's clamp* can be used. The purpose of the clamp is to shunt the current from the
base through diode D0, in dependence on the collector–emitter voltage, to shift the operating point of the transistor from the hard saturation line to the quasi-saturation region (see Figure 2.13). Diodes D1 through D3 produce appropriate bias for the clamping diode D0 (in practice, the number of diodes in the series can be greater or less than three). Diode D4 provides the path for negative base current during turn-off.

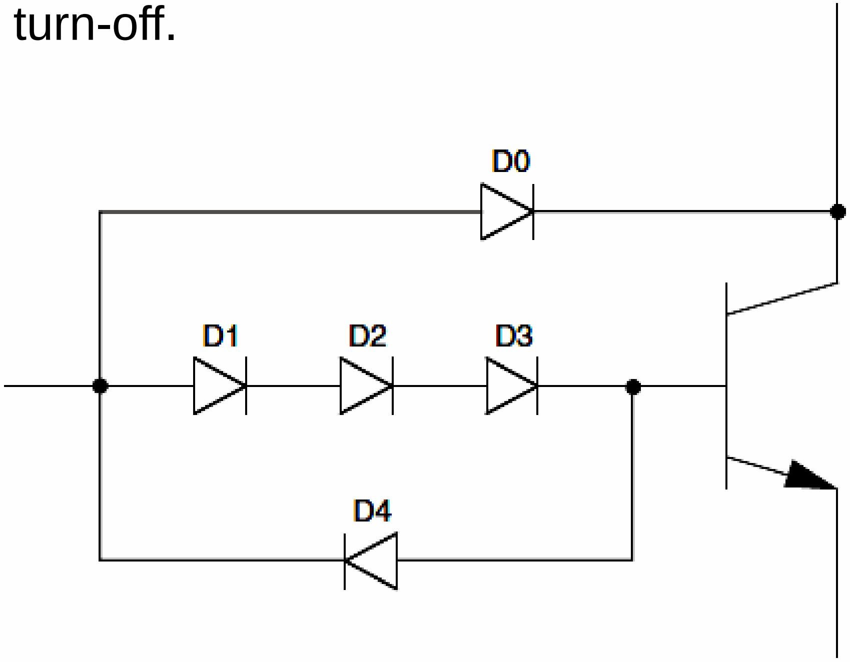


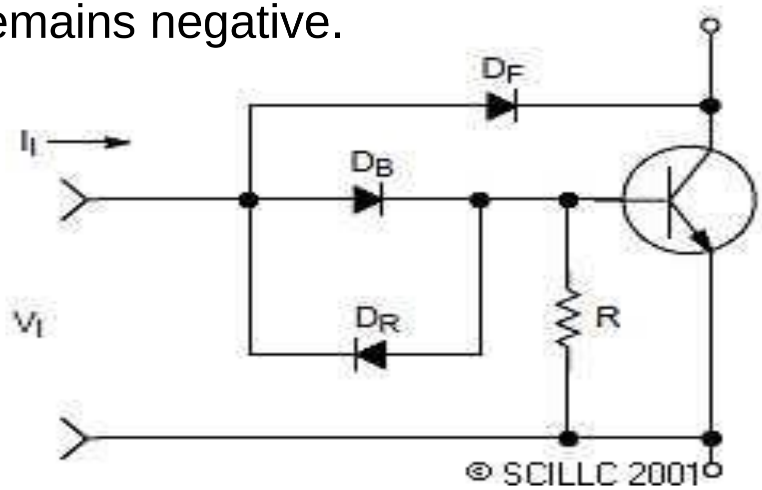
Figure 3.7 Antisaturation Baker's clamp for a BJT.

Baker's Clamp

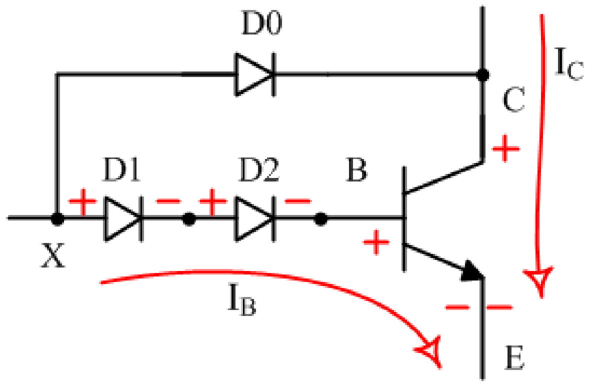
Baker clamp is a generic name for a class of electronic circuits that reduce the storage time of a switching bipolar junction transistor (BJT) by applying a nonlinear negative feedback through various kinds of diodes. The reason for slow turn-off times of saturated BJTs is the stored charge in the base. It must be removed before the transistor will turn off since the storage time is a limiting factor of using bipolar transistors and IGBTs in fast switching applications. The diode-based Baker clamps prevent the transistor from saturating and thereby accumulating a lot of stored charge.[1]

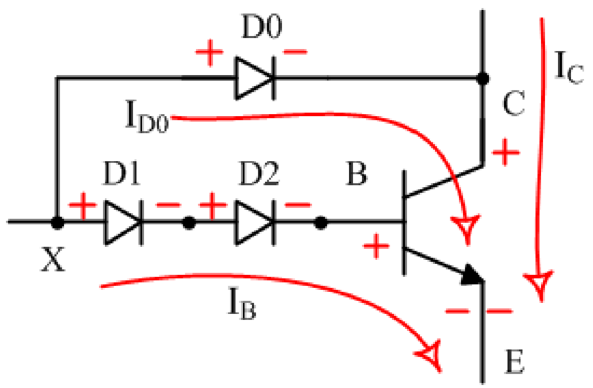
Baker's Clamp

Circuit operation is as follows. When a positive pulse is applied to an initially off transistor, forward current flows through D_B into the base of the transistor, turning it on. The voltage at the input is set at two diode drops, approximately 1.5 V. As collector current builds up and collector voltage drops, diode D_F starts to conduct and keeps V_{CE} from falling below about 0.8 V. This action prevents the bipolar operating point from entering the hard saturation region and theoretically eliminates storage time since the collector-base junction is not reverse biased. Diode D_F acts as a current-regulating feedback element, setting the base current to the amount required by the transistor at a particular collector current and on-state voltage. Current at the input which exceeds this base current (i.e., excess current) is shunted into the collector circuit by diode D_F. When the input pulse goes negative, reverse current is drawn from the base through diode D_R in parallel with diode D_B until its stored charge is depleted. Resistor R is present mainly for noise immunity; it also establishes a small reverse off-state voltage at the base if the input level remains negative.



Baker's Clamp





Let us assume that all diodes have similar characteristics, i.e.,

D0 = D1 = D2

As current I_B flows, then

$$V_{XE} = V_{D1} + V_{D2} + V_{BE} \quad \text{.........(1)}$$

Also, as current I_{D0} flow we have,

$$V_{XE} = V_{XC} + V_{CE} \quad \text{....................(2)}$$

And,

$$V_{XC} = V_{D0}$$

or

$$V_{XE} = V_{D0} + V_{CE} \quad \text{....................(3)}$$

Equating (1) and (3)

$$V_{D0} + V_{CE} = V_{D1} + V_{D2} + V_{BE}$$

$$V_{CE} = V_{D1} + V_{D2} + V_{BE} - V_{D0}$$

When diodes are in forward baised mode, we have

$$V_{D0} = V_{D1} = V_{D2}$$

Therefore,

$$V_{CE} = V_{D1} + V_{BE}$$

And if V_{D1} and V_{BE} are both equal, which is possible because both are PN junctions, hence

$$V_{CE} = 2V_{BE} \approx 1.5V$$