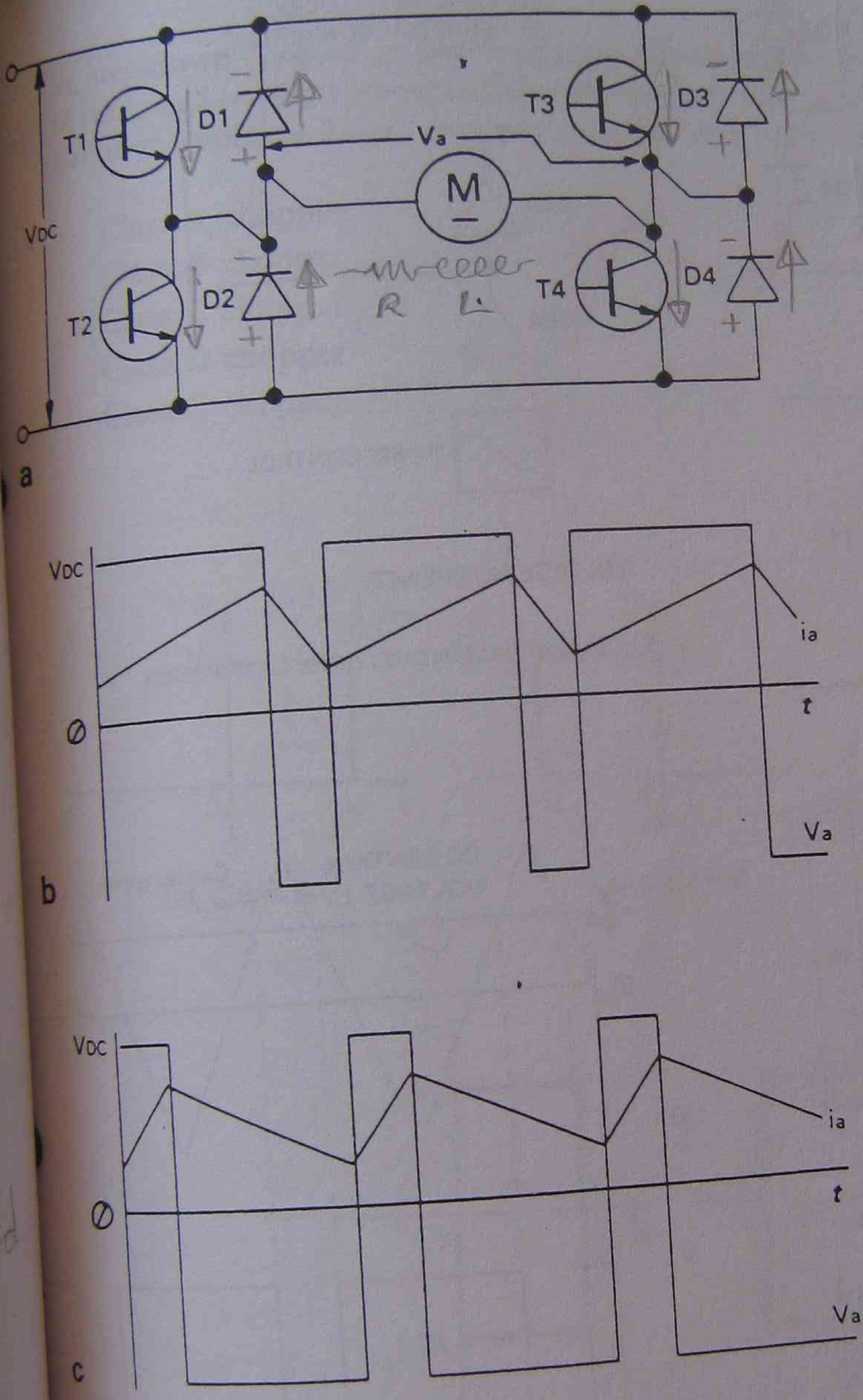
A.C. Motor Control.

Topic -



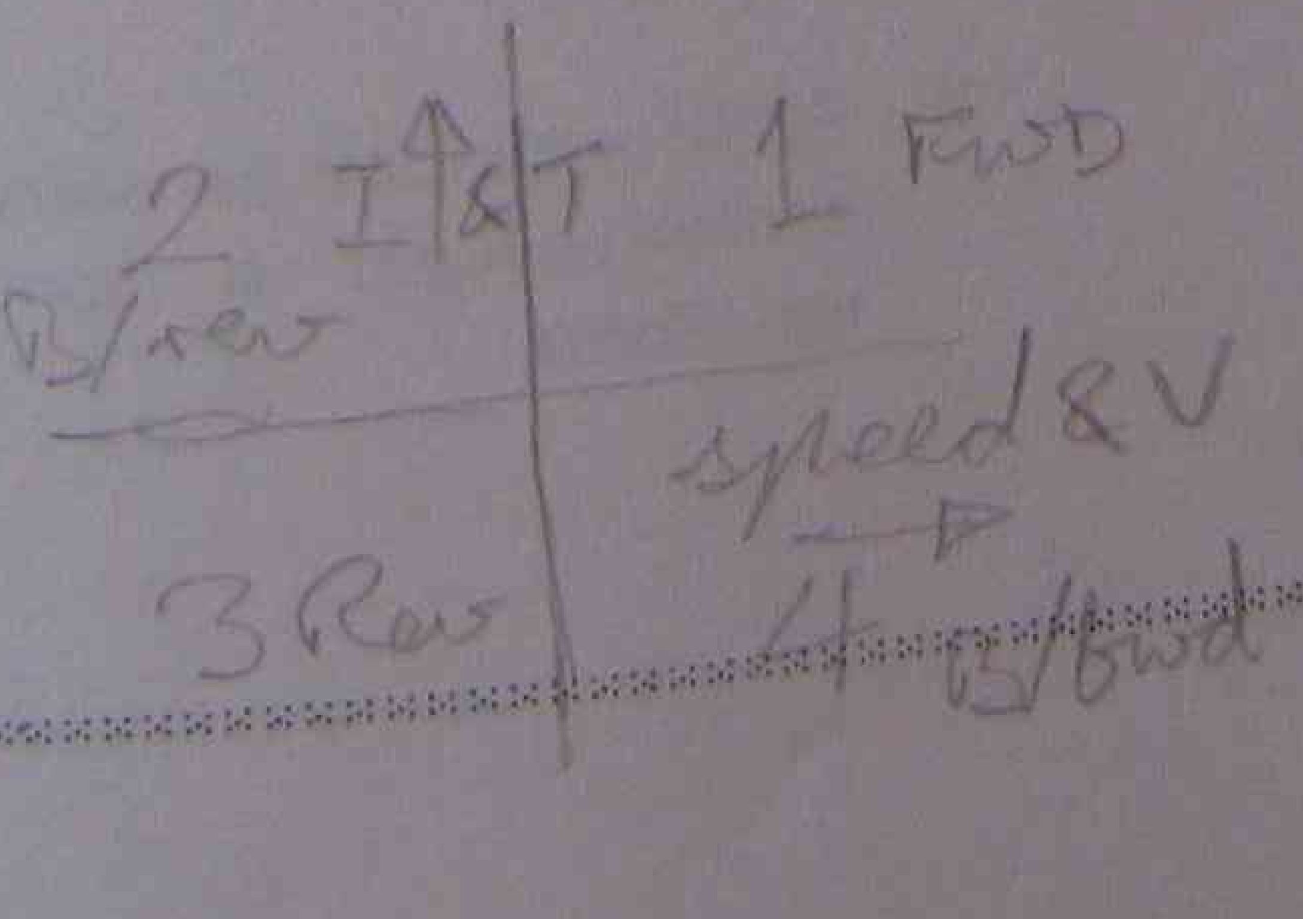
4 quadrant operation

Forward & Reverse
- Motoring &
- Braking.

Forward Motoring

Reverse Braking

4 Four-quadrant dc dc converter; **a** circuit; **b** forward motoring; **c** reverse braking.



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Session No. 3·1

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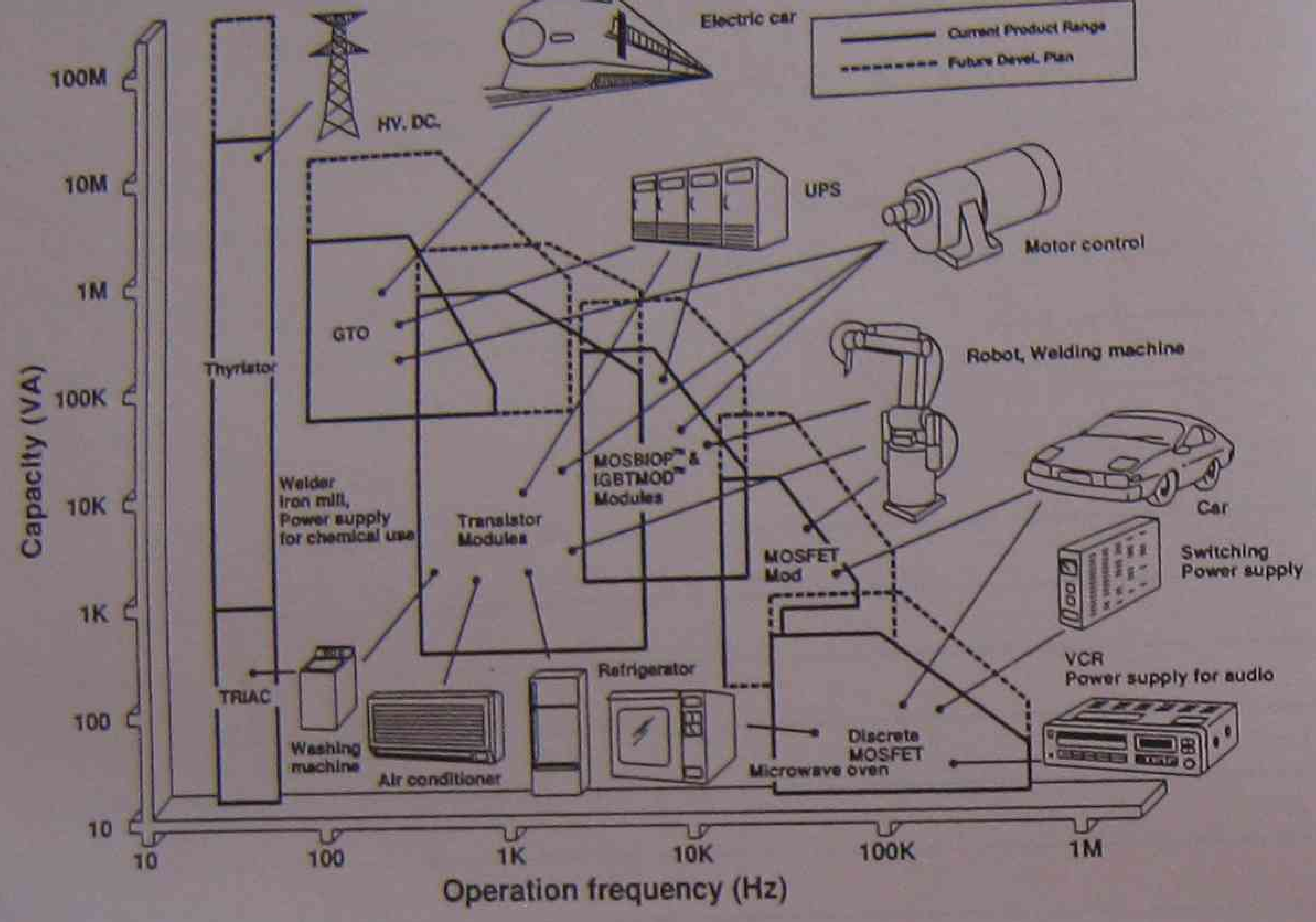
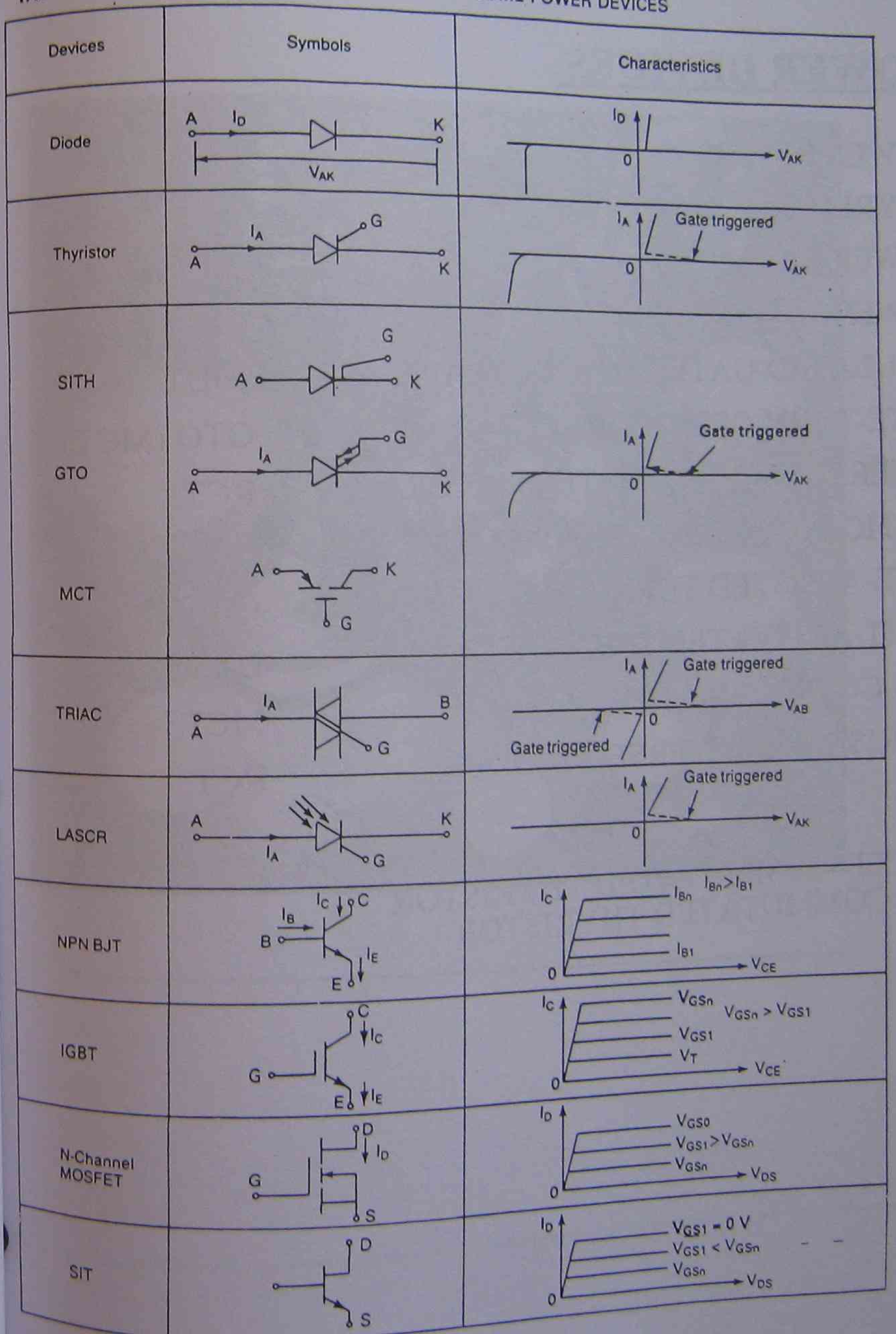


Figure 1-6 Applications of power devices. (Courtesy of Powerex, Inc.)

TABLE 1.2 RATINGS OF POWER SEMICONDUCTOR DEVICES

Type		Voltage/current rating	Upper frequency (Hz)	Switching time (μs)	On-state resistance (Ω)
Diodes	General purpose	5000 V/5000 A	1k	100	0.16m
	High speed	3000 V/1000 A	10k	2–5	1m
	Schottky	40 V/60 A	20k	0.23	10m
Forced-turned-off thyristors	Reverse blocking	5000 V/5000 A	1k	200	0.25m
	High speed	1200 V/1500 A	10k	20	0.47m
	Reverse blocking	2500 V/400 A	5k	40	2.16m
	Reverse conducting	2500 V/1000 A	5k	40	2.1m
	GATT	1200 V/400 A	20k	8	2.24m
	Light triggered	6000 V/1500 A	400	200–400	0.53m
TRIACs		1200 V/300 A	400	200–400	3.57m
Self-turned-off thyristors	GTO	4500 V/3000 A	10k	15	2.5m
	SITH	4000 V/2200 A	20k	6.5	5.75m
Power transistors	Single	400 V/250 A	20k	9	4m
		400 V/40 A	20k	6	31m
		630 V/50 A	25k	1.7	15m
	Darlington	1200 V/400 A	10k	30	10m
SITs		1200 V/300 A	100k	0.55	1.2
Power MOSFETS	Single	500 V/8.6 A	100k	0.7	0.6
		1000 V/4.7 A	100k	0.9	2
		500 V/50 A	100k	0.6	0.4m
IGBTs	Single	1200 V/400 A	20k	2.3	60m
MCTs	Single	600 V/60 A	20k	2.2	18m

TABLE 1.3 CHARACTERISTICS AND SYMBOLS OF SOME POWER DEVICES



Devices	Symbols	Characteristics
Diode	A I_D K V_{AK}	I_D 0 V_{AK}
Thyristor	I_A G A K	I_A Gate triggered 0 V_{AK}
SITH	G A K	
GTO	I_A G A K	I_A Gate triggered 0 V_{AK}
MCT	A K G	
TRIAC	I_A B A G	I_A Gate triggered V_{AB} 0 Gate triggered
LASCR	A K I_A G	I_A Gate triggered V_{AK} 0
NPN BJT	I_C C I_B B I_E E	I_C I_{Bn} $I_{Bn} > I_{B1}$ I_{B1} 0 V_{CE}
IGBT	C I_C G E I_E	I_C V_{GSn} $V_{GSn} > V_{GS1}$ V_{GS1} V_T 0 V_{CE}
N-Channel MOSFET	D I_D G S	I_D V_{GS0} $V_{GS1} > V_{GSn}$ V_{GSn} 0 V_{DS}
SIT	D S	I_D $V_{GS1} = 0$ V $V_{GS1} < V_{GSn}$ V_{GSn} 0 V_{DS}

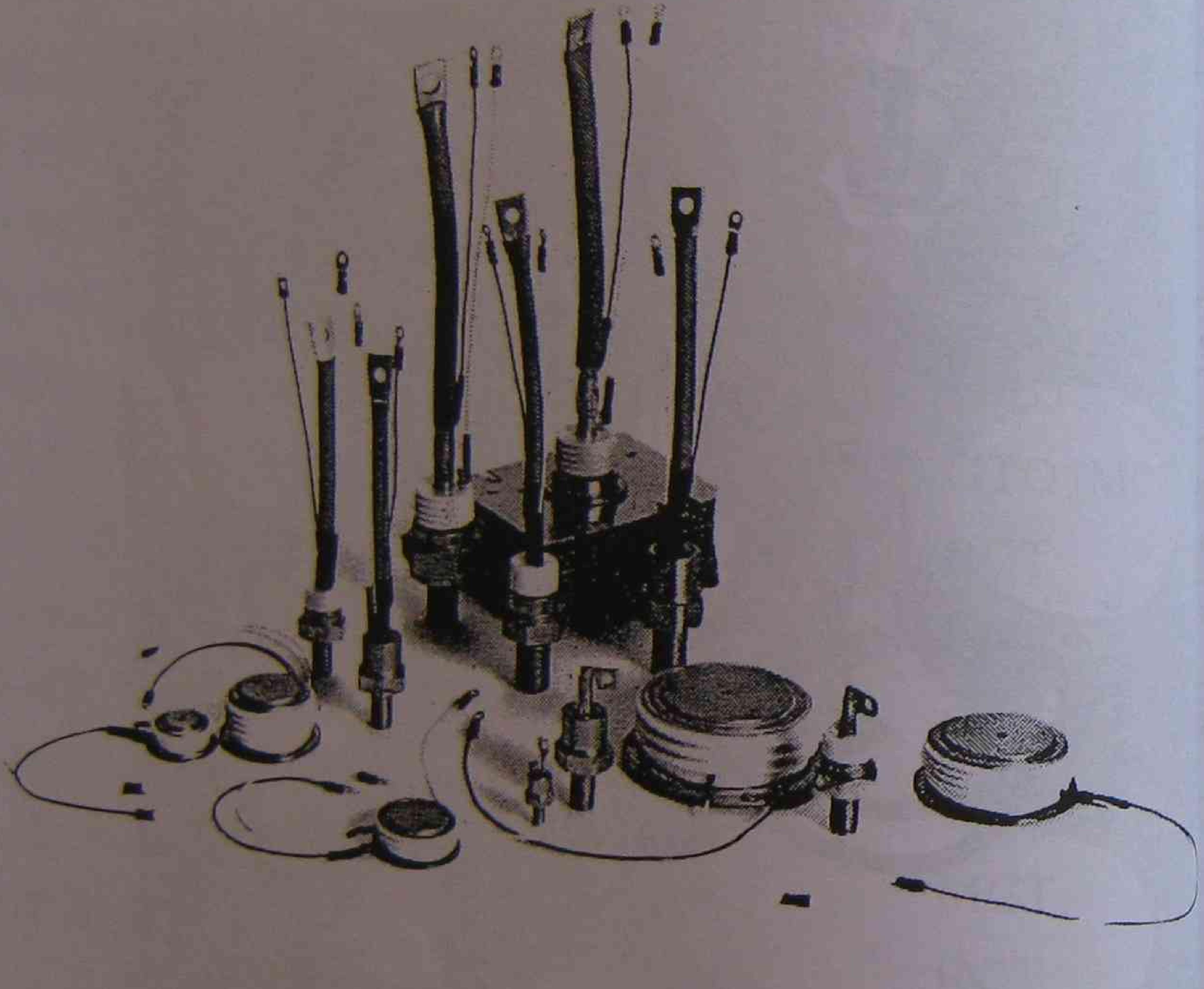
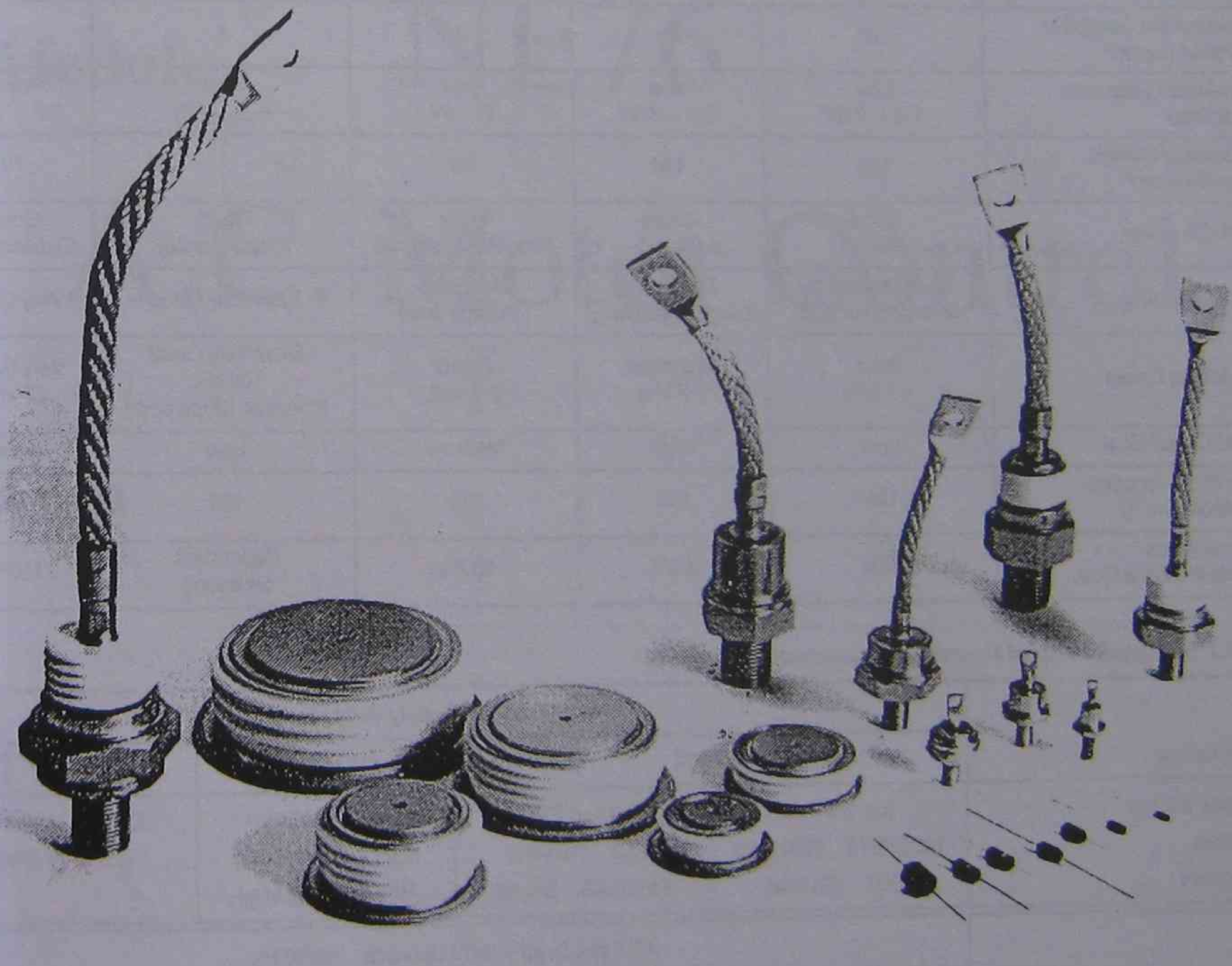


Figure 1-3 Various general-purpose diode configurations. (Courtesy of Powerex, Inc.)



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A.C. Motor Control

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Session No. 3·2

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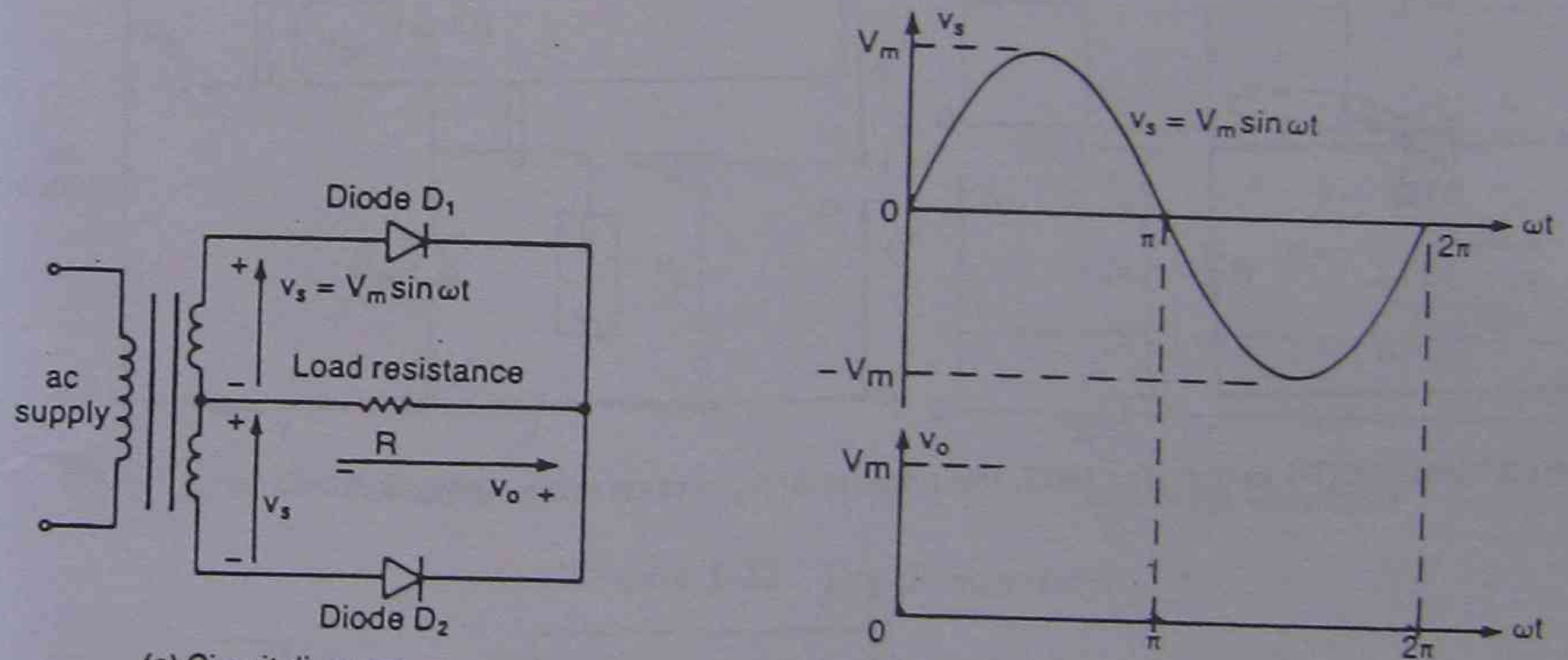
Objectives :-

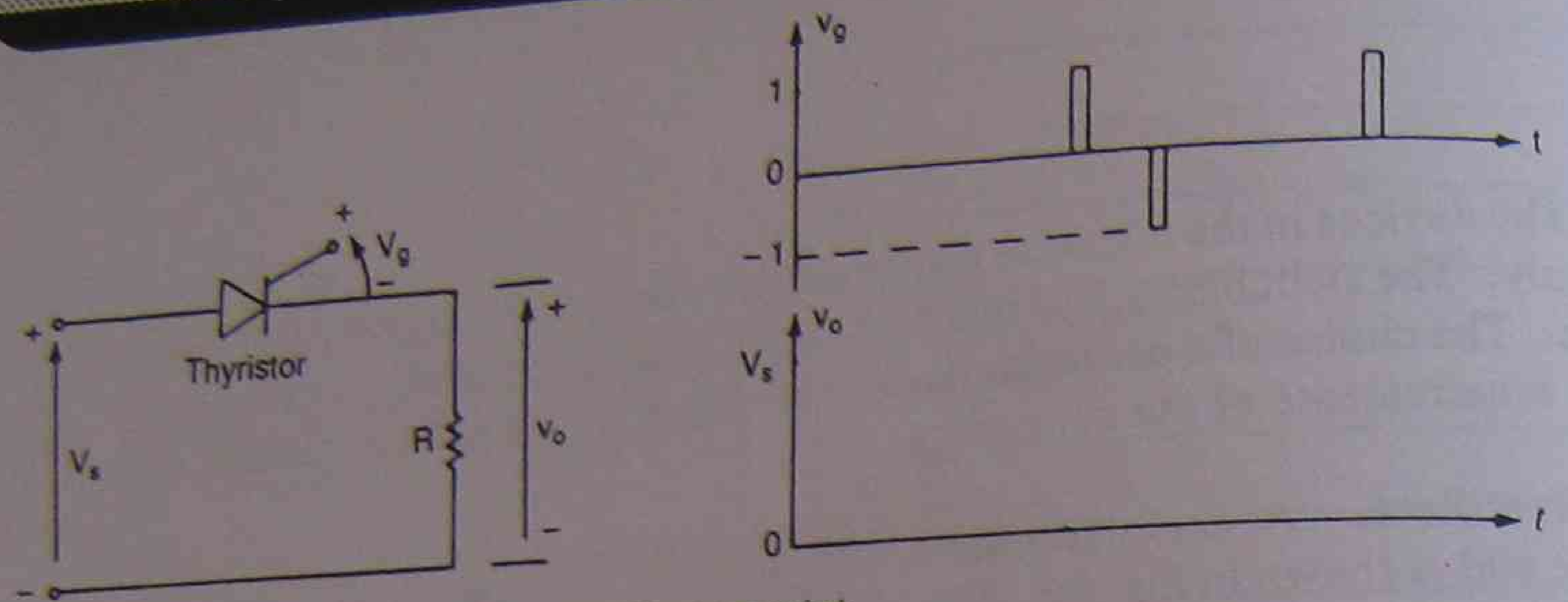
The devices in the following converters are used to illustrate the basic principles only. The switching action of a converter can be performed by more than one device. The choice of a particular device will depend on the voltage, current, and speed requirements of the converter.

Rectifiers. A diode rectifier circuit converts ac voltage into a fixed dc voltage and is shown in Fig. 1-8. The input voltage to the rectifier could be either single-phase or three-phase.

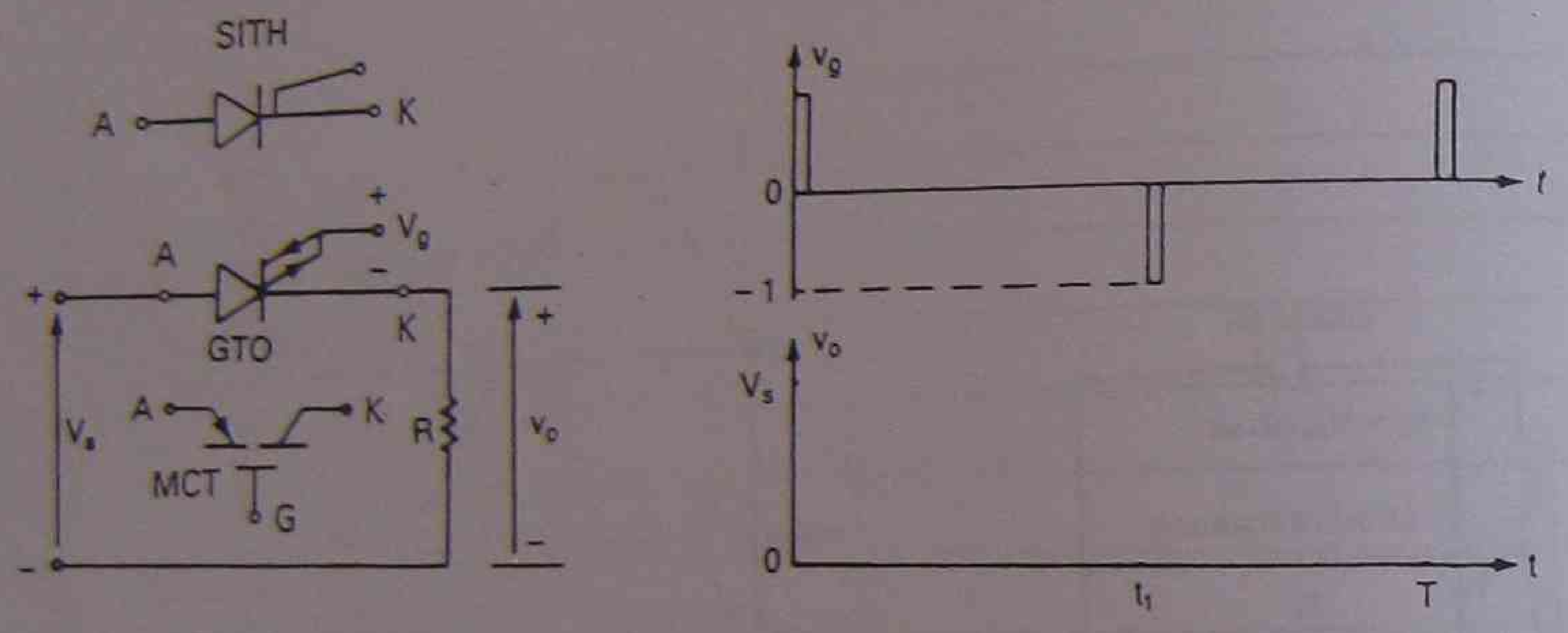


(a) Circuit diagram (b) Voltage waveforms

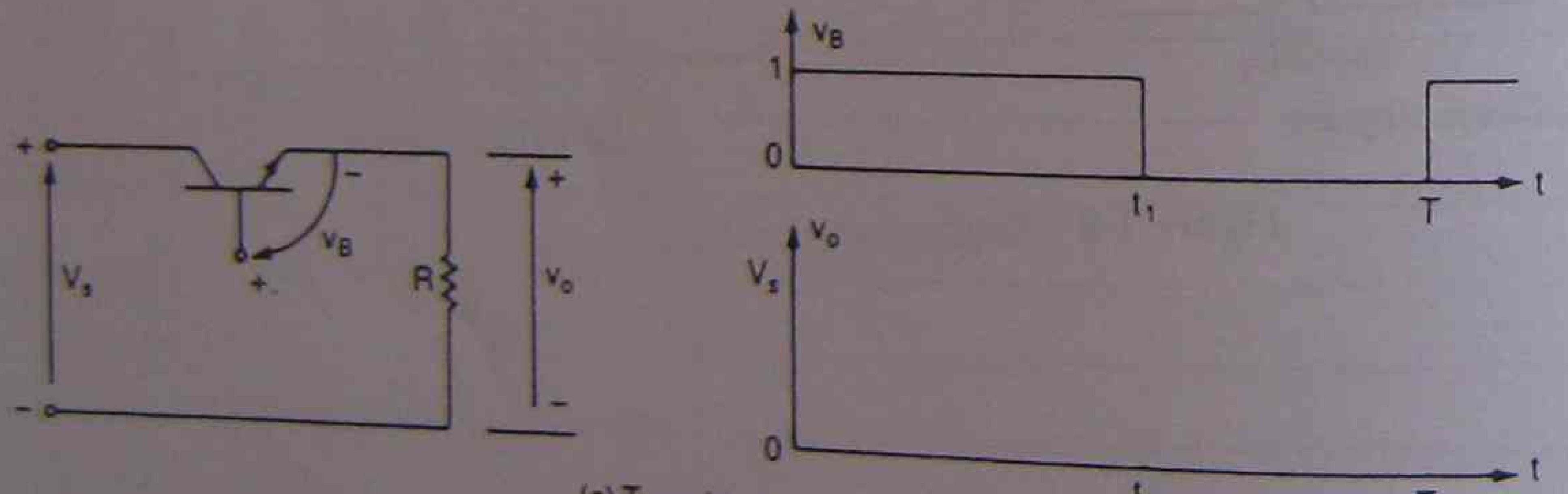
Figure 1-8 Single-phase rectifier circuit.



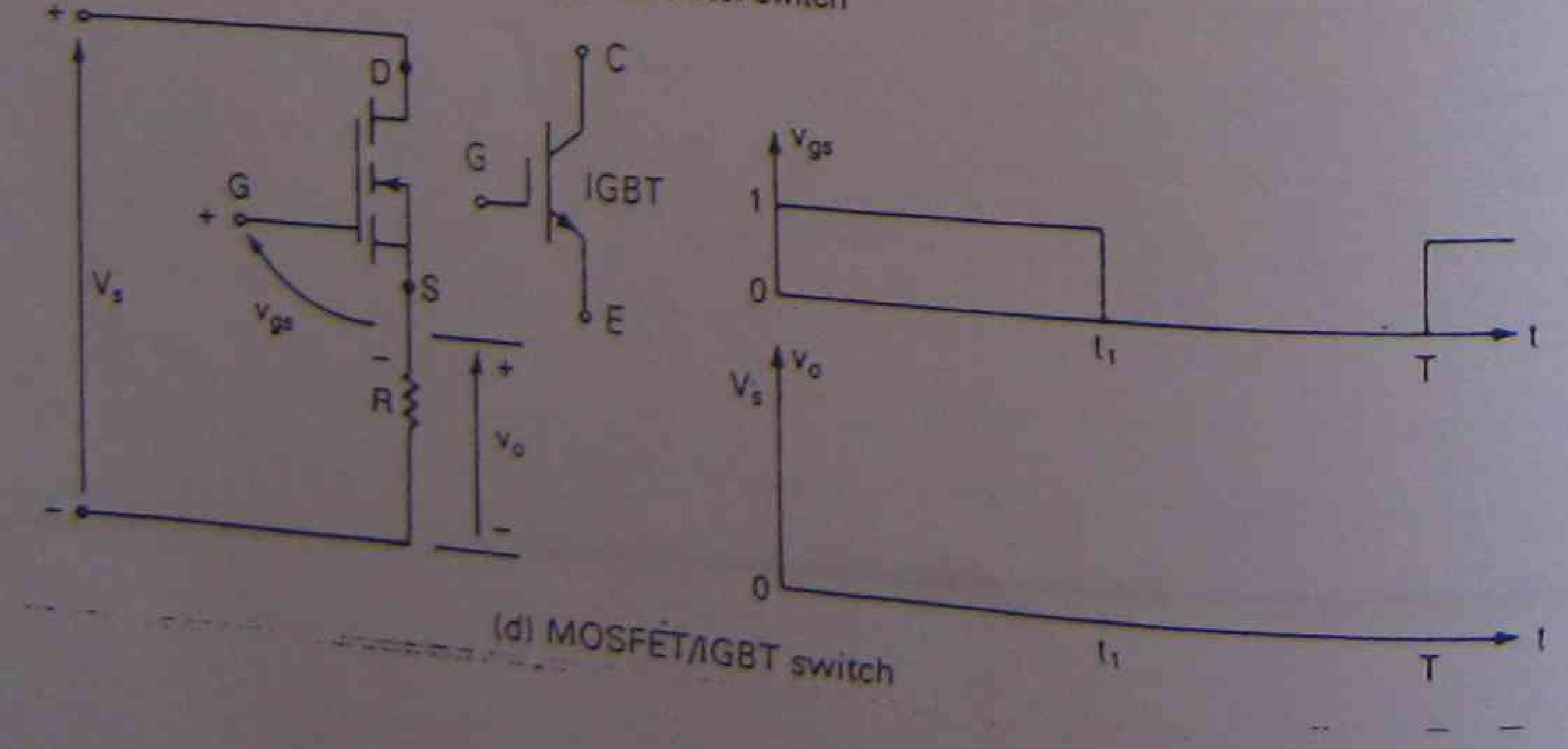
(a) Thyristor switch



(b) GTO/MCT/SITH switch (For MCT, the polarity of V_g is reversed as shown

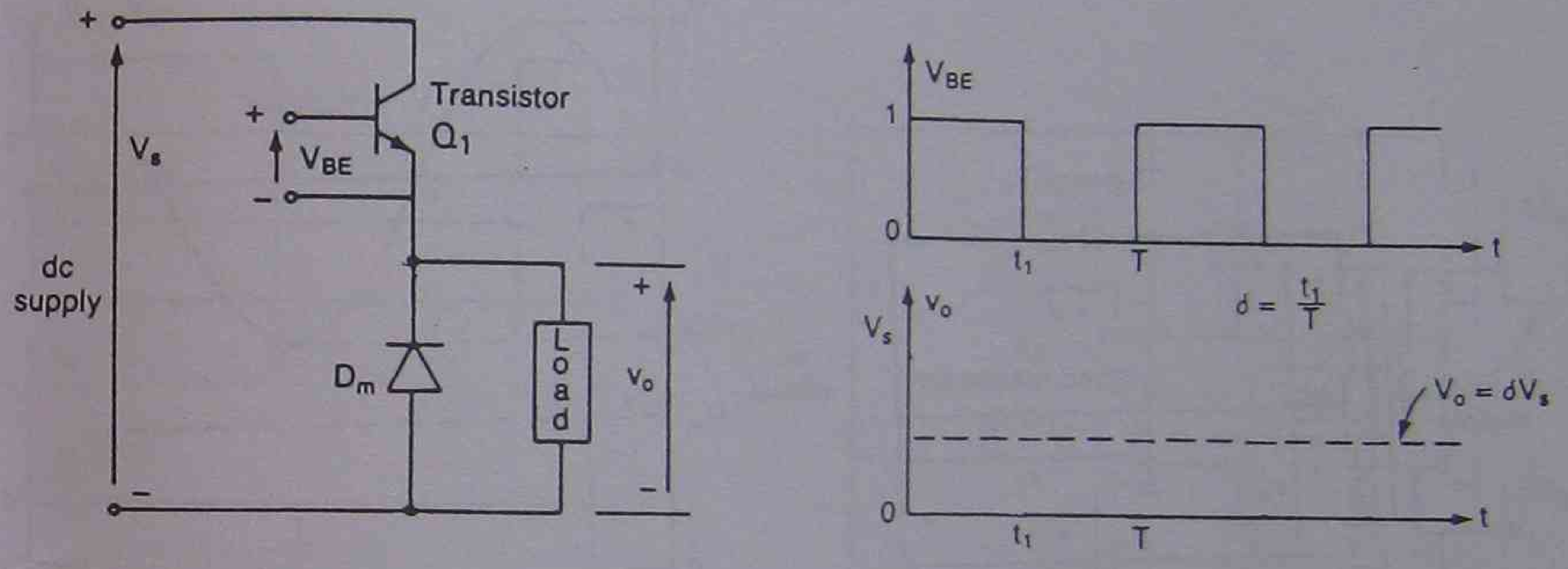


(c) Transistor switch



(d) MOSFET/IGBT switch

Dc–dc converters. A dc–dc converter is also known as a *chopper* or *switching regulator* and a transistor chopper is shown in Fig. 1-11. The average output voltage is controlled by varying the conduction time t_1 of transistor Q_1. If T is the chopping period, then $t_1 = \delta T$. δ is called as the *duty cycle* of the chopper.

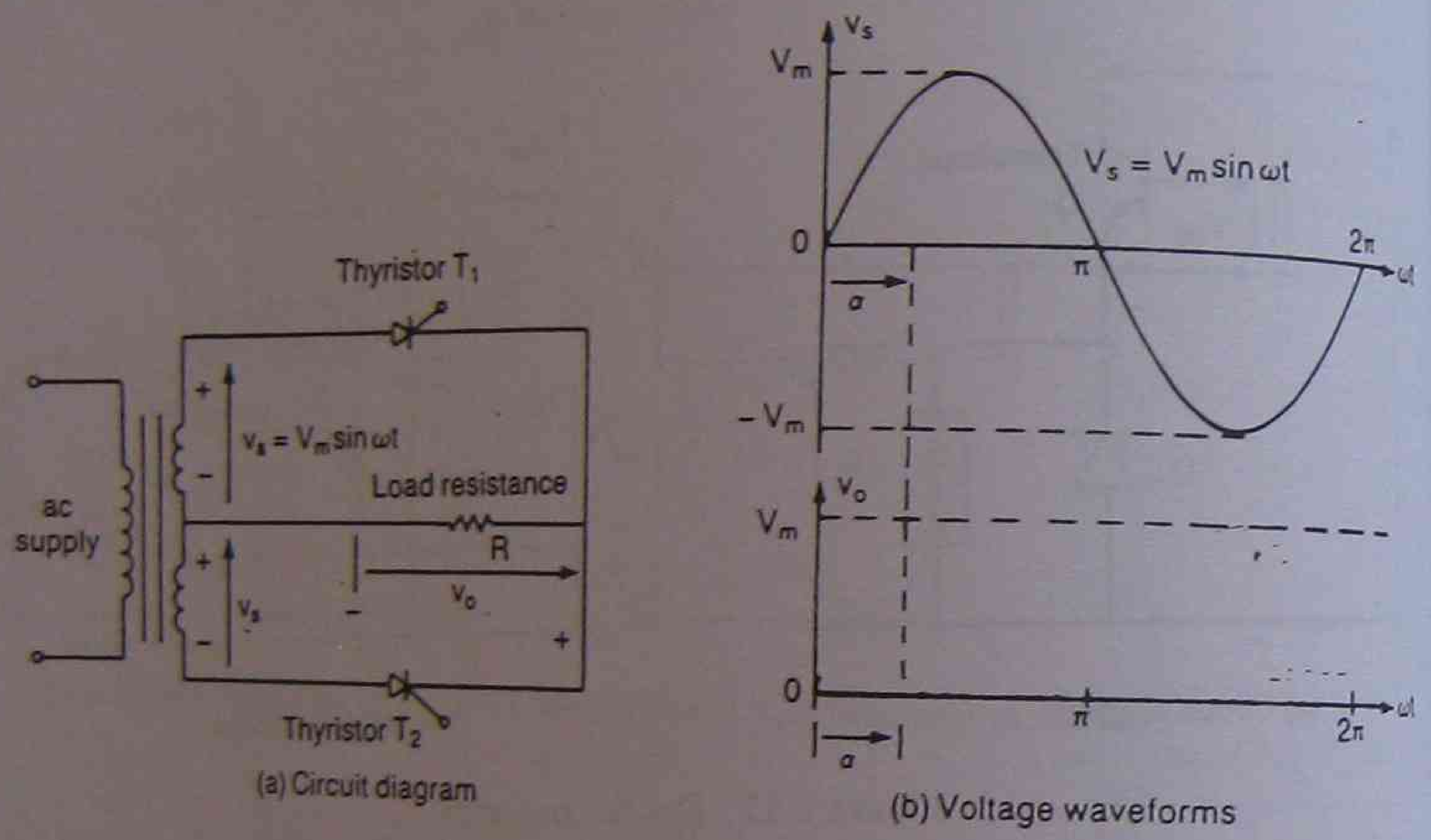


(a) Circuit diagram

(b) Voltage waveforms

Figure 1-11 Dc–dc converter.

Ac–dc converters. A single-phase converter with two natural commutated thyristors is shown in Fig. 1-9. The average value of the output voltage can be controlled by varying the conduction time of thyristors or firing delay angle, α. The input could be a single or three-phase source. These converters are als known as *controlled rectifiers*.

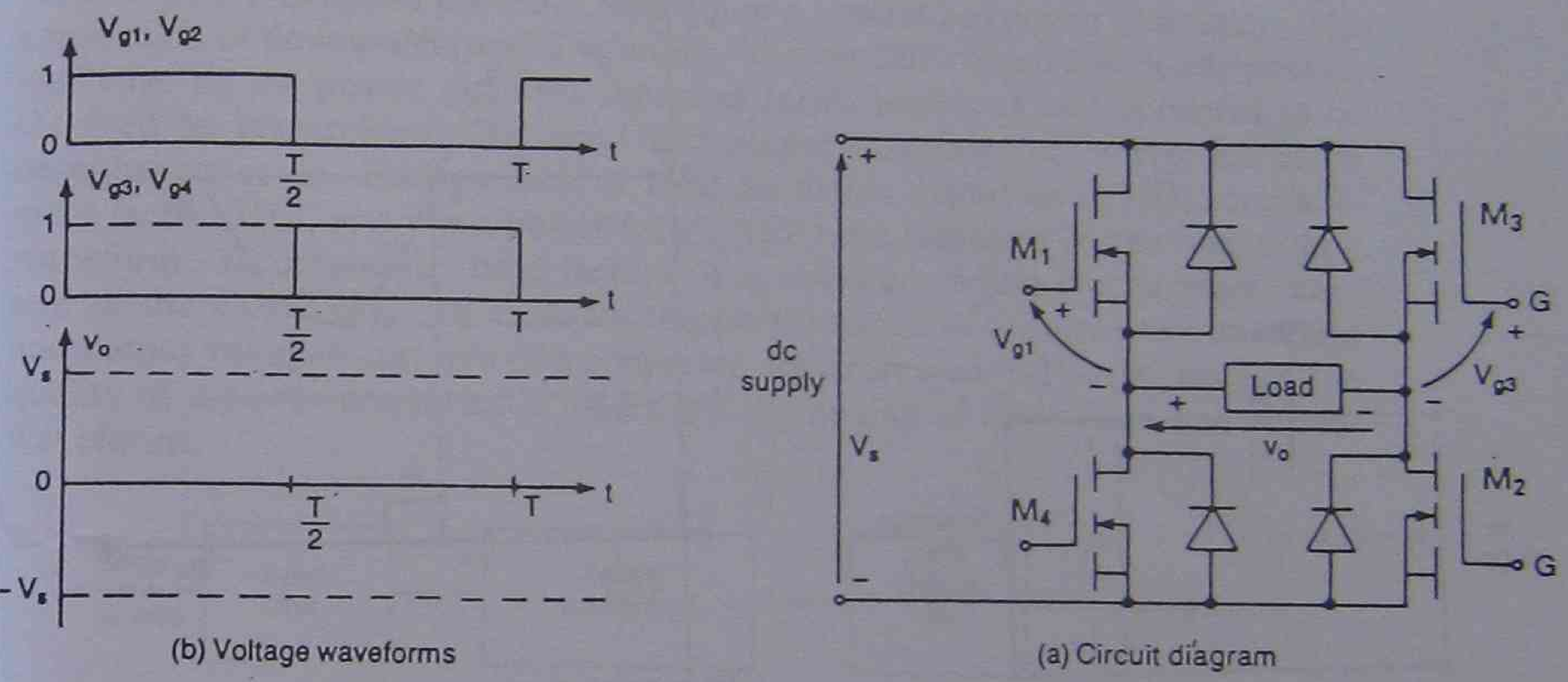


(a) Circuit diagram

(b) Voltage waveforms

Figure 1-9 Single-phase ac–dc converter.

Dc–ac converters. A dc–ac converter is also known as an *inverter*. A single-phase transistor inverter is shown in Fig. 1-12. If transistors M_1 and M_2 conduct for one-half period and M_3 and M_4 conduct for the other half, the output voltage is of alternating form. The output voltage can be controlled by varying the conduction time of transistors.

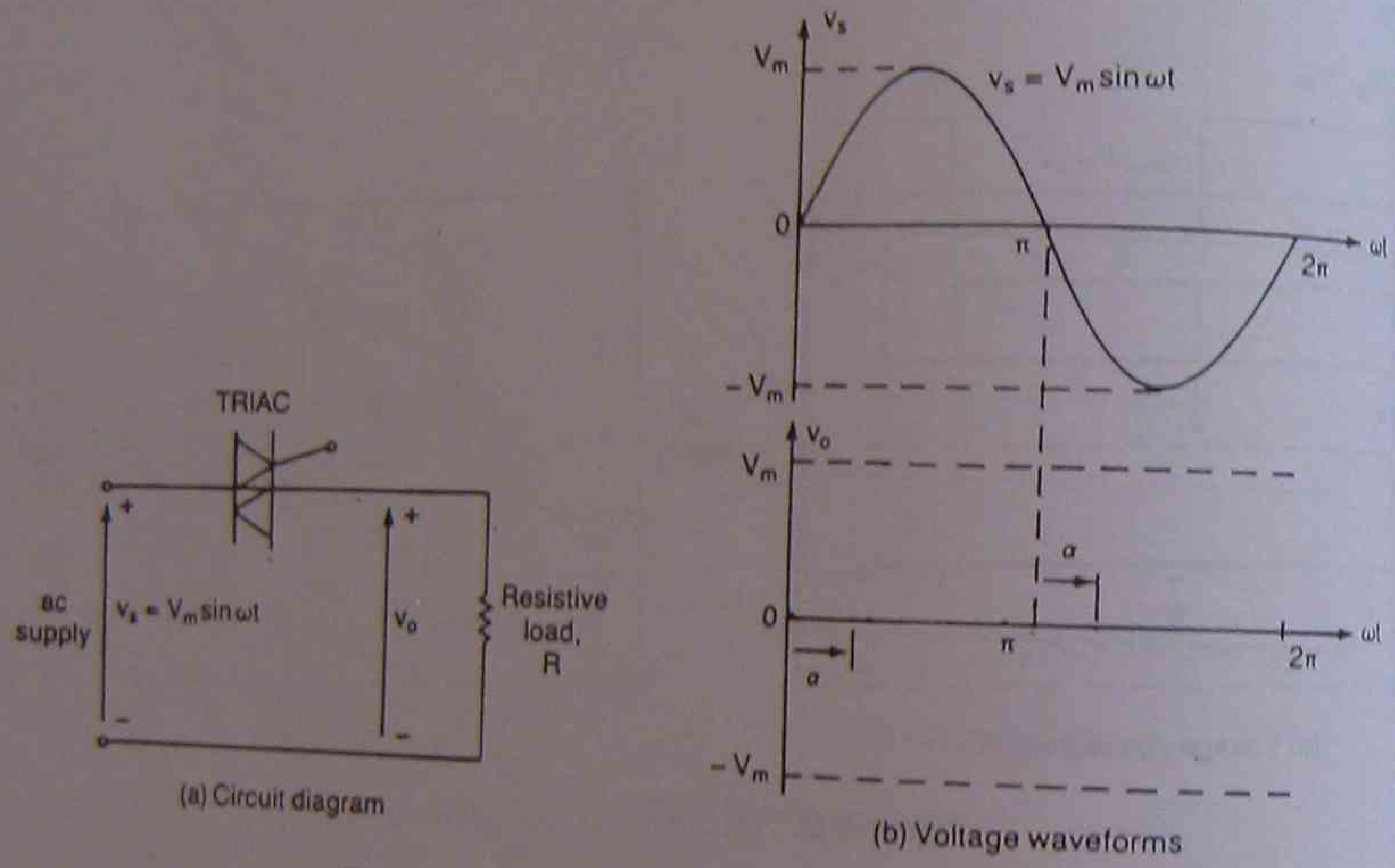


(b) Voltage waveforms

(a) Circuit diagram

Figure 1-12 Single-phase dc–ac converter.

Ac–ac converters. These converters are used to obtain a variable ac output voltage from a fixed ac source and a single-phase converter with a TRIAC is shown in Fig. 1-10. The output voltage is controlled by varying the conduction time of a TRIAC or firing delay angle, α. These types of converters are also known as *ac voltage controllers*.



(a) Circuit diagram

(b) Voltage waveforms

Figure 1-10 Single-phase ac–ac converter.

The operations of the power converters are based mainly on the switching of power semiconductor devices; and as a result the converters introduce current and voltage harmonics into the supply system and on the output of the con- verters. These can cause problems of distortion of the output voltage, harmonic generation into the supply system, and interference with the communication and signaling circuits. It is normally necessary to introduce filters on the input and output of a converter system to reduce the harmonic level to an acceptable magni- tude. Figure 1-13 shows the block diagram of a generalized power converter. The application of power electronics to supply the sensitive electronic loads poses a challenge on the power quality issues and raises problems and concerns to be resolved by researchers. The input and output quantities of converters could be either ac or dc. Factors such as total harmonic distortion (THD), displace- ment factor (HF), and input power factor (IPF) are measures of the quality of a waveform. To determine these factors, it is required to find the harmonic con- tent of the waveforms. To evaluate the performance of a converter, the input and output voltages/currents of a converter are expressed in Fourier series. The quality of a power converter is judged by the quality of its voltage and current waveforms.

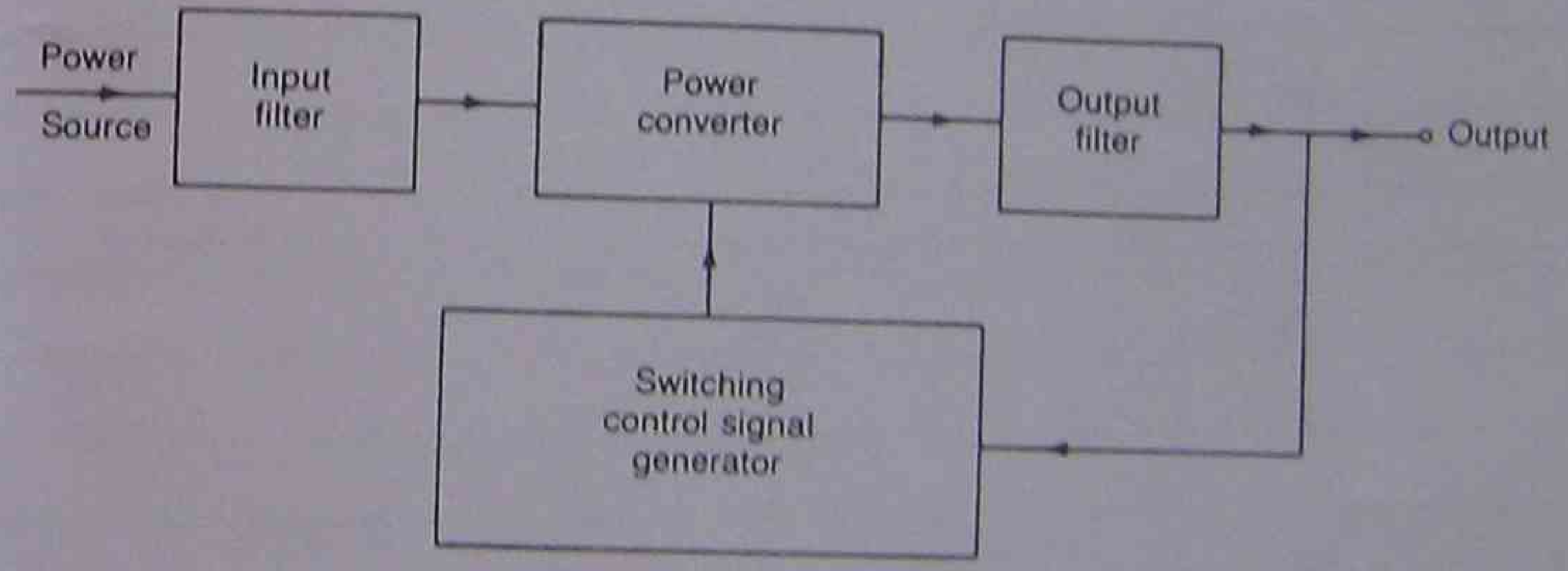


Figure 1-13 Generalized power converter system.

The control strategy for the power converters plays an important part on the harmonic generation and output waveform distortion, and can be aimed to minimize or reduce these problems. The power converters can cause radio-frequency interference due to electromagnetic radiation and the gating circuits may generate erroneous signals. This interference can be avoided by *grounded shielding*.

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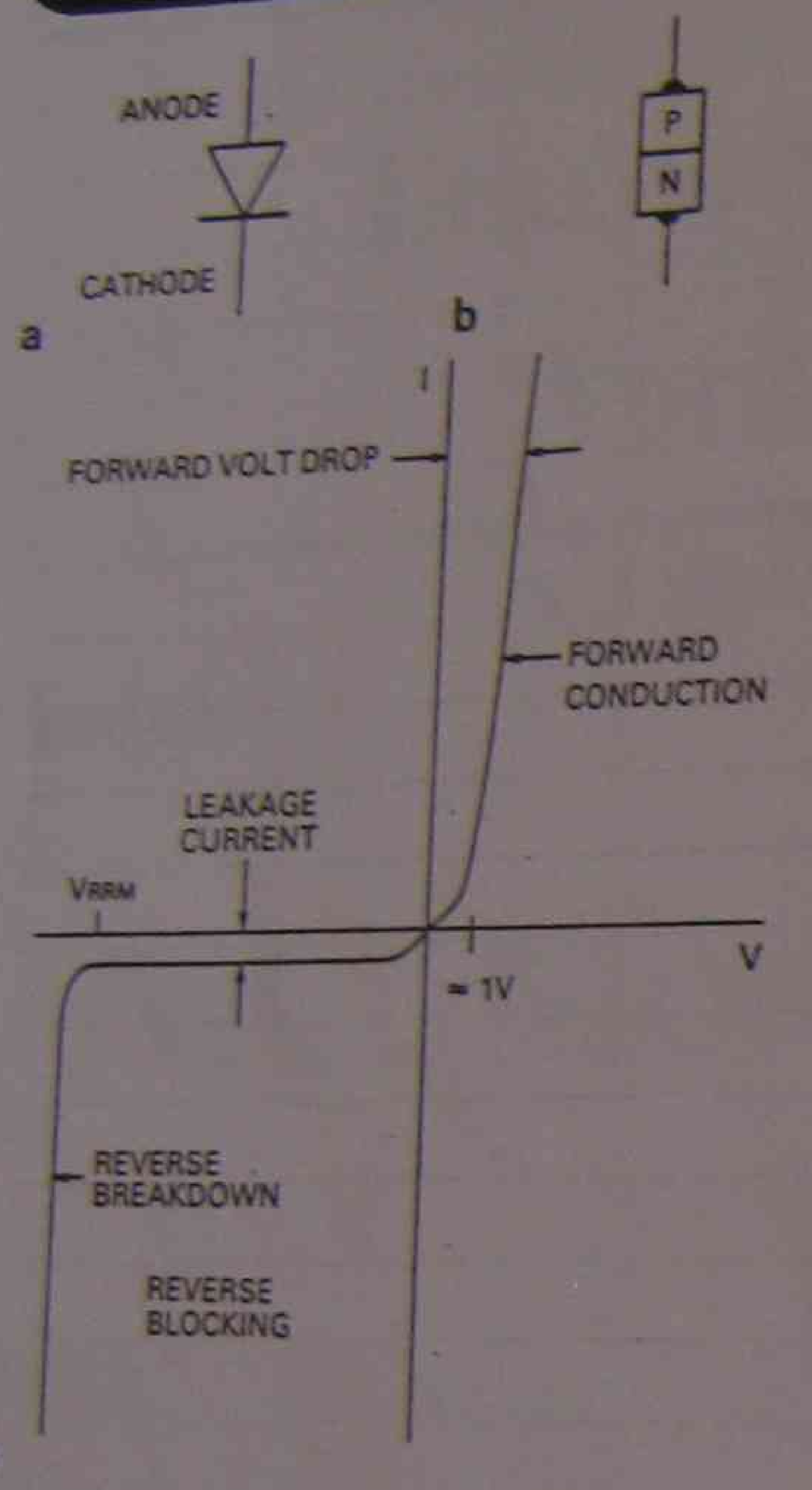
A.C. Motor Control

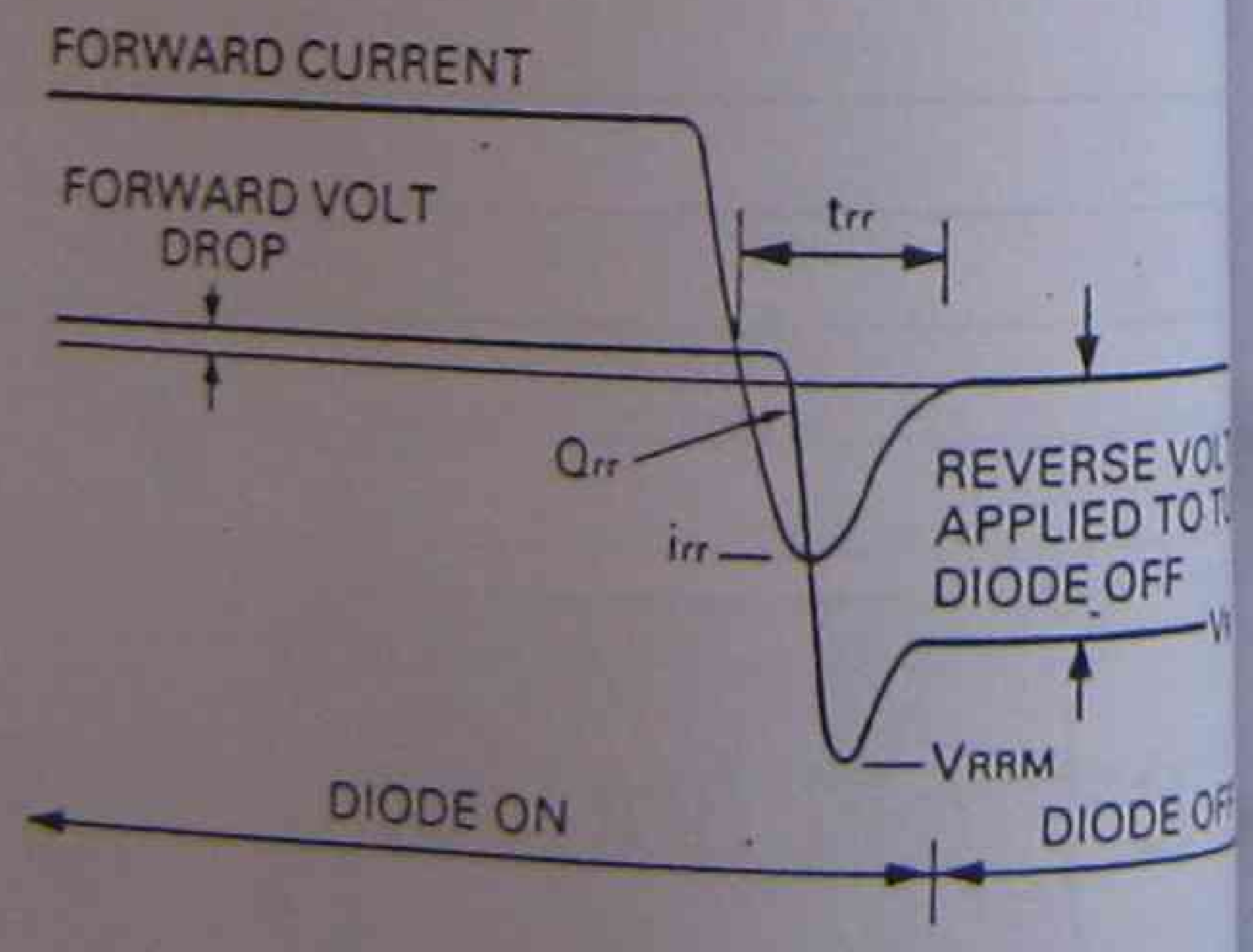
Area ....

Topic ....

Session No. 3



1 *Diode:* **a** *circuit symbol;* **b** *structure;* **c** *characteristic.*



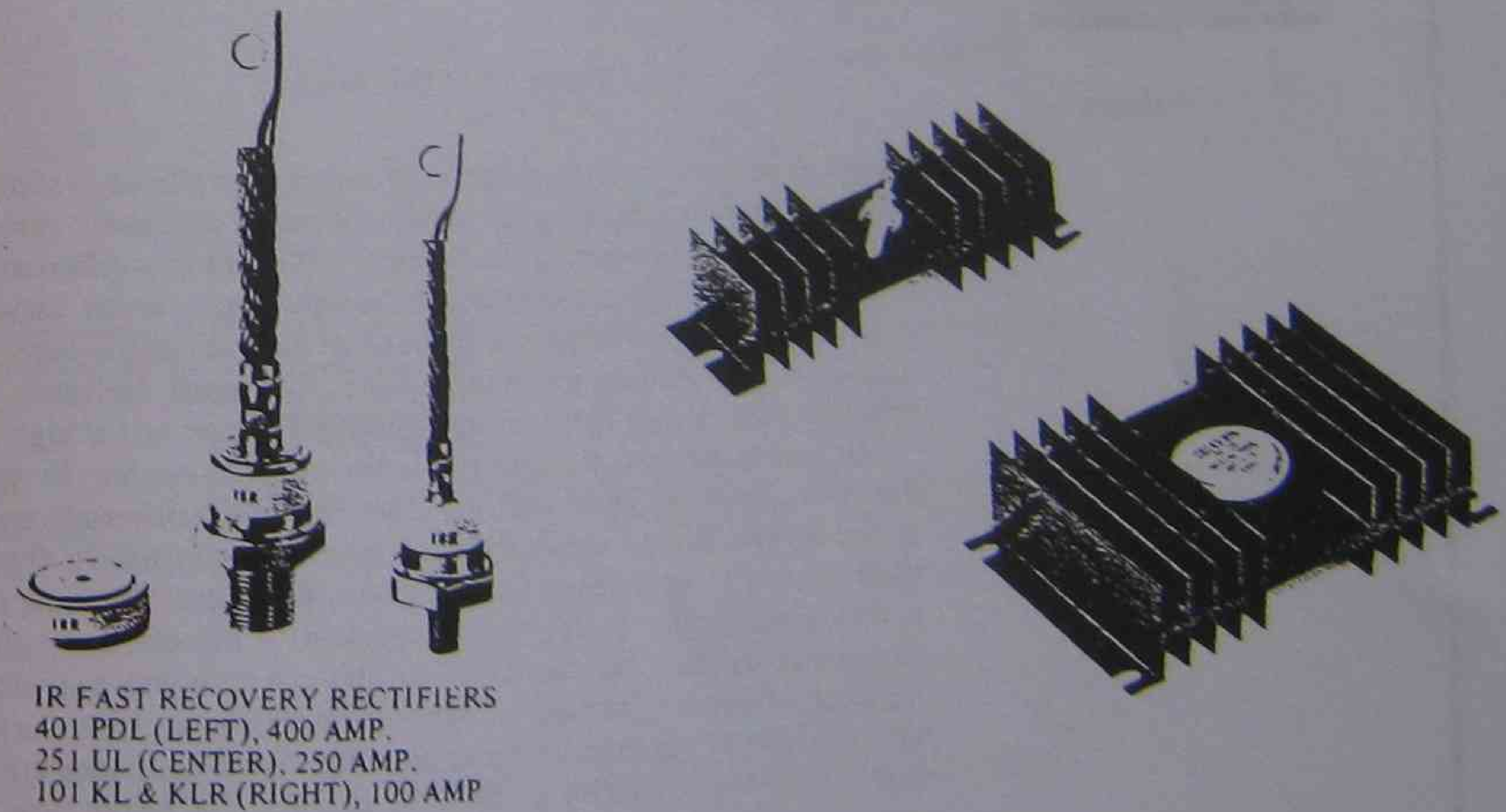
2 *Switching characteristics of a diode*

20.4 POWER DIODES

There are a number of diodes designed specifically to handle the high-power and high-temperature demands of some applications. The most frequent use of power diodes occurs in the rectification process, in which ac signals (having zero average value) are converted to ones having an average or dc level. As noted in Chapter 2, when used in this capacity, diodes are normally referred to as *rectifiers*.

The majority of the power diodes are constructed using silicon because of its higher current, temperature, and PIV ratings. The higher current demands require that the junction area be larger, to ensure that there is a low forward diode resistance. If the forward resistance were too large, the I^2R losses would be excessive. The current capability of power diodes can be increased by placing two or more in parallel and the PIV rating can be increased by stacking the diodes in series.

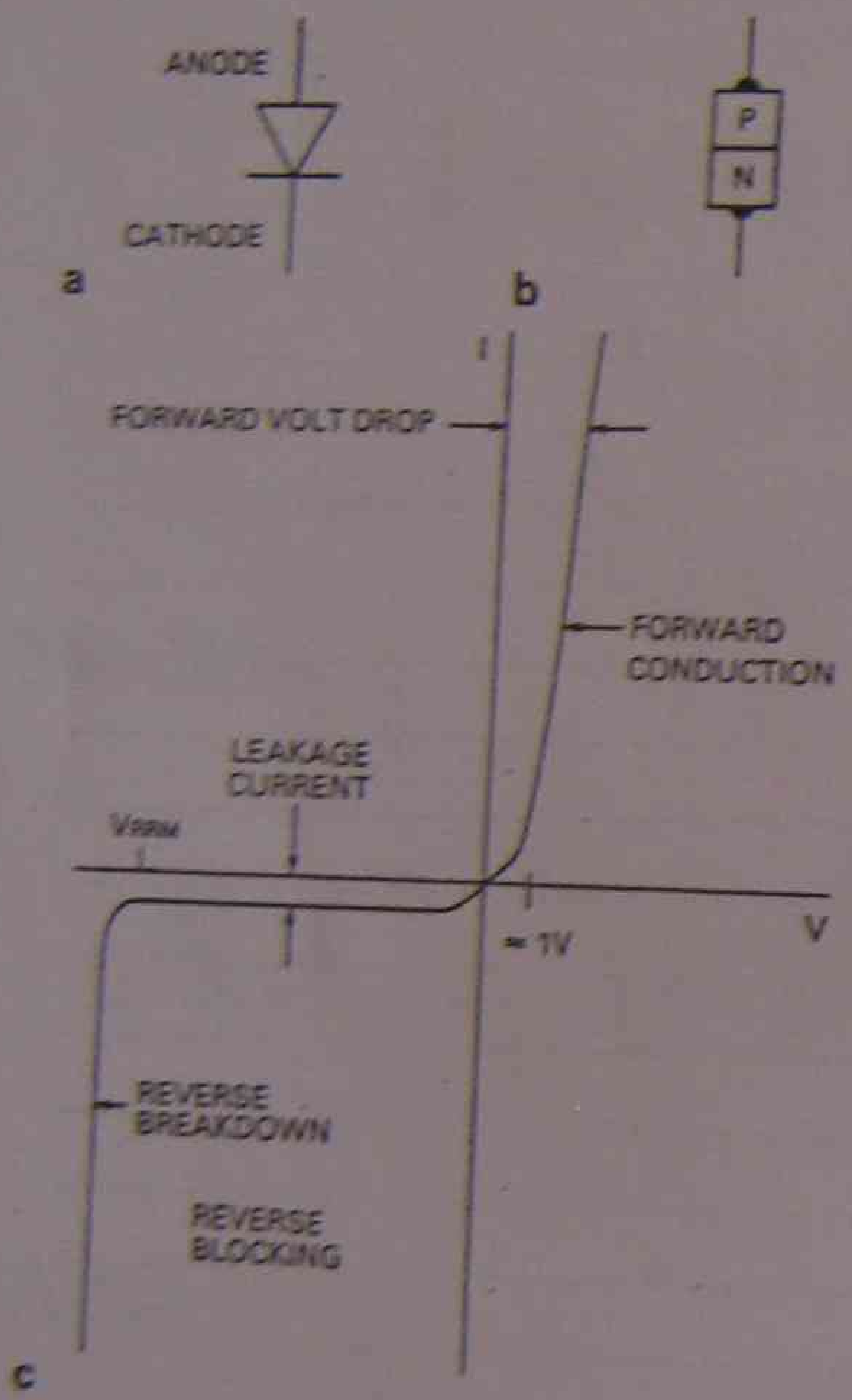
Various types of power diodes and their current rating have been provided in Fig. 20.12a. The high temperatures resulting from the heavy current require, in many cases, that heat sinks be used to draw the heat away from the element. A few of the various types of heat sinks available are shown in Fig. 20.12b. If heat sinks are not employed, stud diodes are designed to be attached directly to the chassis, which in turn will act as the heat sink.



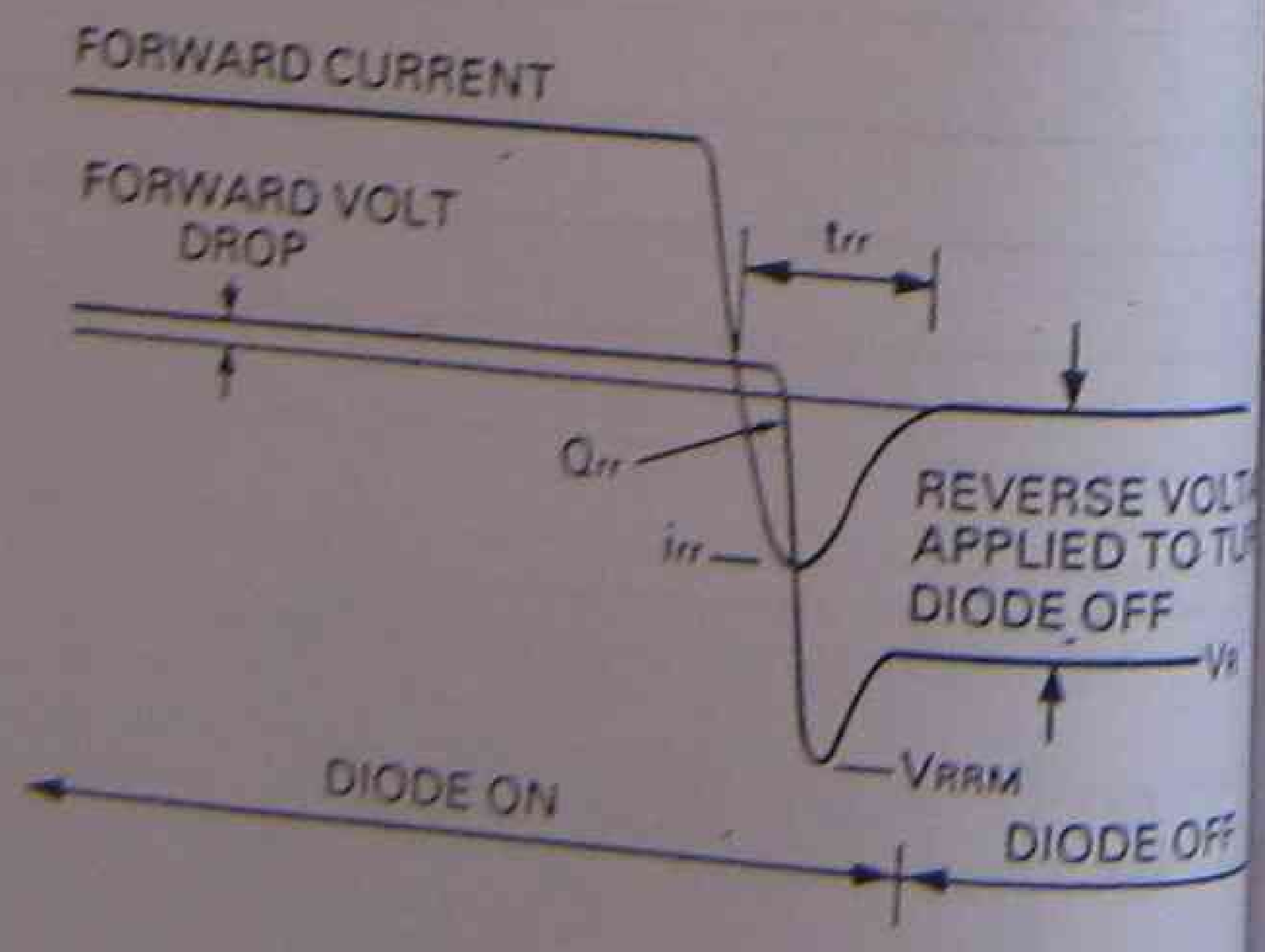
IR FAST RECOVERY RECTIFIERS
401 PDL (LEFT), 400 AMP.
251 UL (CENTER), 250 AMP.
101 KL & KLR (RIGHT), 100 AMP

(a) (b)

Figure 20.12 Power diodes and heat sinks. (Courtesy International Rectifier Corporation.)



1 *Diode:* **a** *circuit symbol;* **b** *structure;* **c** *characteristic.*



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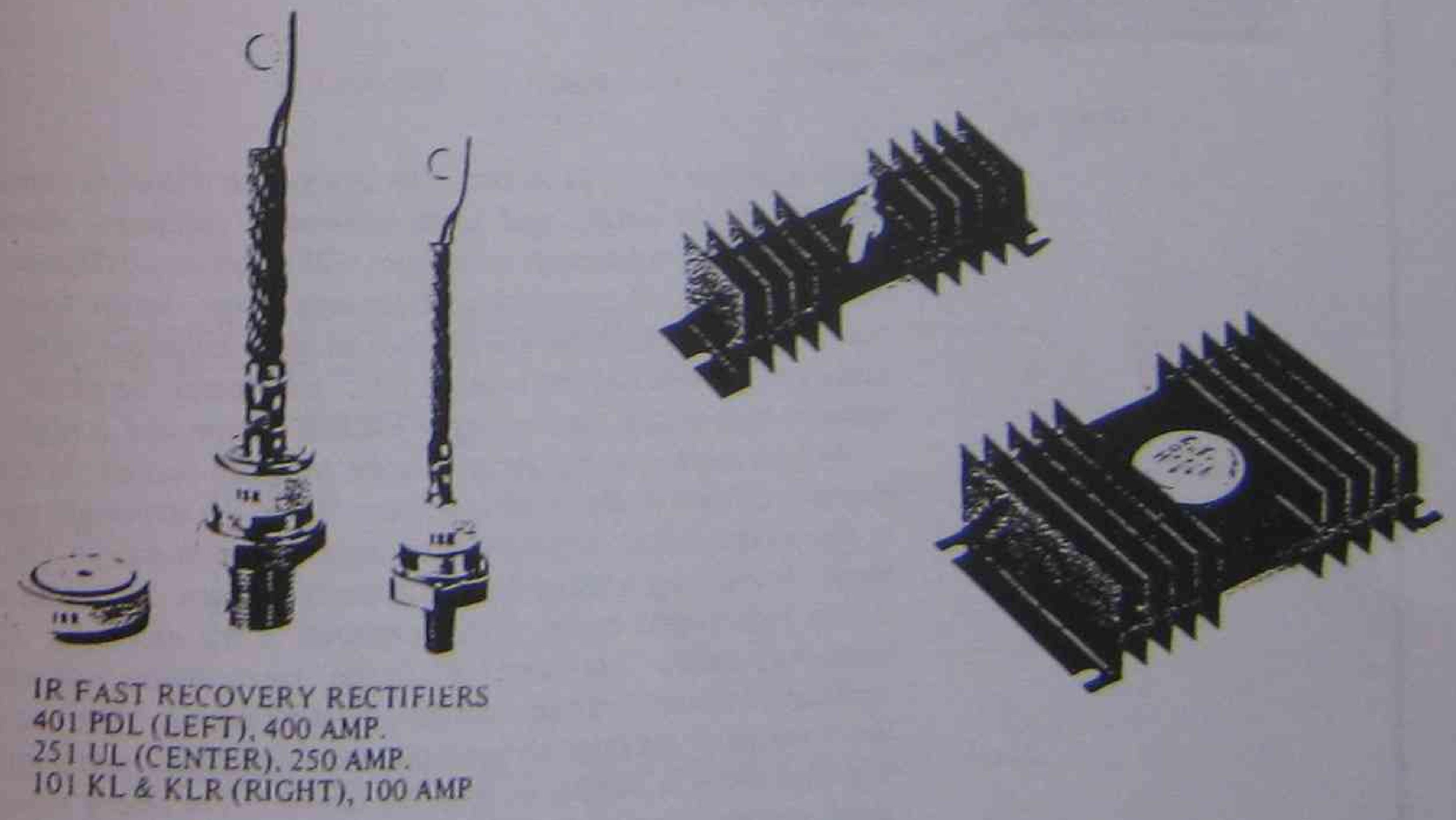


(a) (b)

Figure 20.12 Power diodes and heat sinks. (Courtesy International Rectifier Corporation.)

20.2 SCHOTTKY BARRIER (HOT-CARRIER) DIODES

In recent years there has been increasing interest in a two-terminal device referred to as a *Schottky-barrier*, *surface-barrier*, or *hot-carrier* diode. Its areas of application were first limited to the very high frequency range due to its quick response time (especially important at high frequencies) and a lower noise figure (a quantity of real importance in high-frequency applications). In recent years, however, it is appearing more and more in low-voltage/high-current power supplies and ac-to-dc converters. Other areas of application of the device include radar systems, Schottky TTL logic for computers, mixers and detectors in communication equipment, instrumentation, and analog-to-digital converters.

Its construction is quite different from the conventional *p-n* junction in that a metal-semiconductor junction is created such as shown in Fig. 20.1. The semiconductor is normally *n*-type silicon (although *p*-type silicon is sometimes used), wh host of different metals, such as molybdenum, platinum, chrome, or tungsten used. Different construction techniques will result in a different set of characteris for the device, such as increased frequency range, lower forward bias, and so Priorities do not permit an examination of each technique here, but information usually be provided by the manufacturer. In general, however, Schottky diode c struction results in a more uniform junction region and a high level of ruggedne

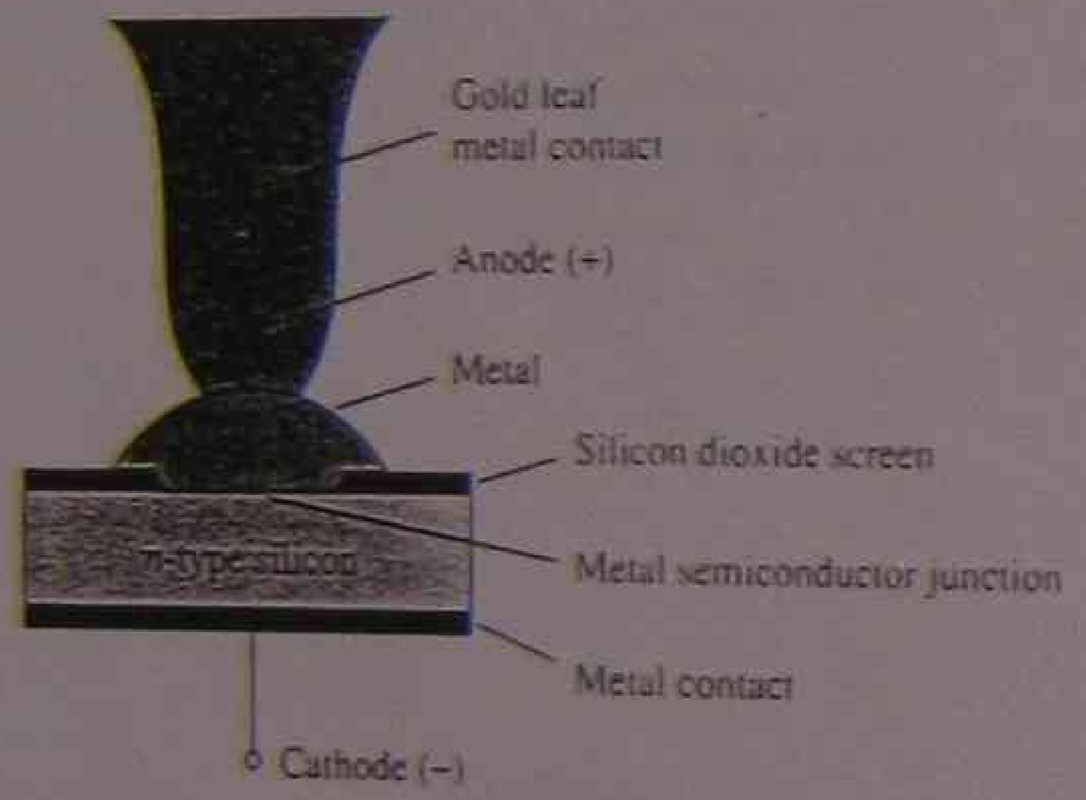


Figure 20.1 Passivated hot-carrier diode

In both materials, the electron is the majority carrier. In the metal, the leve minority carriers (holes) is insignificant. When the materials are joined the elect in the *n*-type silicon semiconductor material immediately flow into the adjoi metal, establishing a heavy flow of majority carriers. Since the injected carriers h a very high kinetic energy level compared to the electrons of the metal, they commonly called "hot carriers." In the conventional *p-n* junction there was injection of minority carriers into the adjoining region. Here the electrons are injec into a region of the same electron plurality. Schottky diodes are therefore unique that conduction is entirely by majority carriers. The heavy flow of electrons into metal creates a region near the junction surface depleted of carriers in the sil material—much like the depletion region in the *p-n* junction diode. The additi carriers in the metal establish a "negative wall" in the metal at the boundary betw the two materials. The net result is a "surface barrier" between the two mater preventing any further current. That is, any electrons (negatively charged) in silicon material face a carrier-free region and a "negative wall" at the surface of metal.

4

The application of a forward bias as shown in the first quadrant of Fig. 20.2 will reduce the strength of the negative barrier through the attraction of the applied positive potential for electrons from this region. The result is a return to the heavy flow of electrons across the boundary, the magnitude of which is controlled by the level of the applied bias potential. The barrier at the junction for a Schottky diode is less than that of the *p-n* junction device in both the forward- and reverse-bias regions. The result is therefore a higher current at the same applied bias in the forward- and reverse-bias regions. This is a desirable effect in the forward-bias region but highly undesirable in the reverse-bias region.

The exponential rise in current with forward bias is described by Eq. (1.4) but with η dependent on the construction technique (1.05 for the metal whisker type of

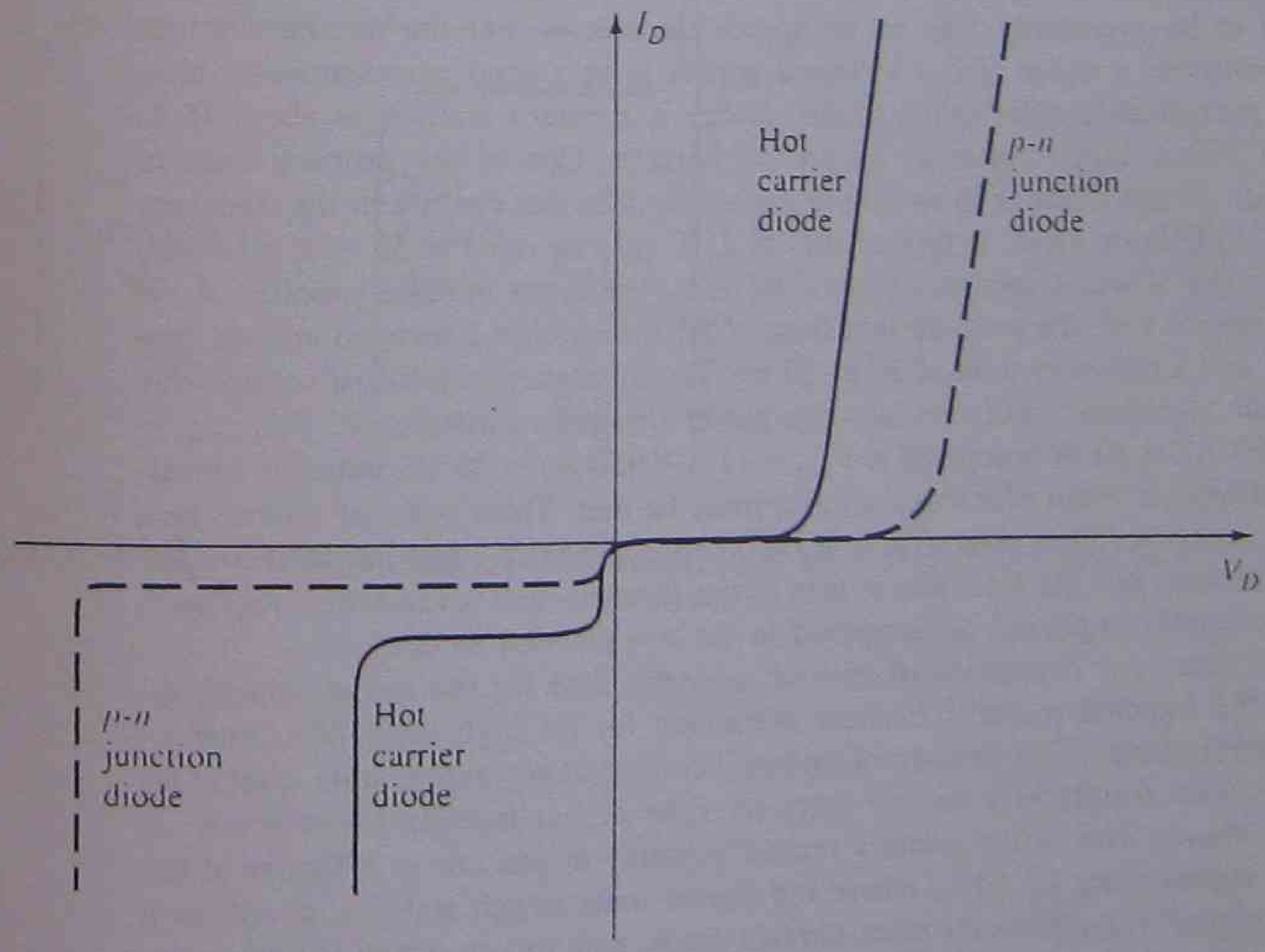


Figure 20.2 Comparison of characteristics of hot-carrier and *p-n* junction diodes.

construction, which is somewhat similar to the germanium diode). In the reverse-bias region the current I_s is due primarily to those electrons in the metal passing into the semiconductor material. One of the areas of continuing research on the Schottky diode centers on reducing the high leakage currents that result with temperatures over 100°C. Through design, improvement units are now becoming available that have a temperature range from −65 to +150°C. At room temperature, I_s is typically in the microampere range for low-power units and milliampere range for high-power devices, although it is typically larger than that encountered using conventional *p-n* junction devices with the same current limits. In addition, the PIV of Schottky diodes is usually significantly less than that of a comparable *p-n* junction unit. Typically, for a 50-A unit, the PIV of the Schottky diode would be about 50 V as compared to 150 V for the *p-n* junction variety. Recent advances, however, have resulted in Schottky diodes with PIVs greater than 100 V at this current level. It is obvious from the characteristics of Fig. 20.2 that the Schottky diode is closer to the ideal set of characteristics than the point contact and has levels of V_T less than the typical silicon semiconductor *p-n* junction. The level of V_T for the "hot-carrier" diode is controlled to a large measure by the metal employed. There exists a required trade-off between temperature range and level of V_T. An increase in one appears to correspond to a resulting increase in the other. In addition, the lower the range of allowable current levels, the lower the value of V_T. For some low-level units, the value of V_T can be assumed to be essentially zero on an approximate basis. For the middle and high range, however, a value of 0.2 V would appear to be a good representative value.

The maximum current rating of the device is presently limited to about 75 A, although 100-A units appear to be on the horizon. One of the primary areas of application of this diode is in *switching power supplies* that operate in the frequency range of 20 kHz or more. A typical unit at 25°C may be rated at 50 A at a forward voltage of 0.6 V with a recovery time of 10 ns for use in one of these supplies. A *p-n* junction device with the same current limit of 50 A may have a forward voltage drop of 1.1 V and a recovery time of 30 to 50 ns. The difference in forward voltage may not appear significant, but consider the power dissipation difference: $P_{\text{hot carrier}} = (0.6\text{ V})(50\text{ A}) = 30\text{ W}$ compared to $P_{p\text{-}n} = (1.1\text{ V})(50\text{ A}) = 55\text{ W}$, which is a measurable difference when efficiency criteria must be met. There will, of course, be a higher dissipation in the reverse-bias region for the Schottky diode due to the higher leakage current, but the total power loss in the forward- and reverse-bias regions is still significantly improved as compared to the *p-n* junction device.

Recall from our discussion of reverse recovery time for the semiconductor di-
ode that the injected minority carriers accounted for the high level of t_{rr} (the re-
verse recovery time). The absence of minority carriers at any appreciable level in the
Schottky diode results in a reverse recovery time of significantly lower levels, as
indicated above. This is the primary reason Schottky diodes are so effective at fre-
quencies approaching 20 GHz, where the device must switch states at a very high
rate. For higher frequencies the point-contact diode, with its very small junction area,
is still employed.

The equivalent circuit for the device (with typical values) and a commonly used
symbol appear in Fig. 20.3. A number of manufacturers prefer to use the standard
diode symbol for the device, since its function is essentially the same. The inductance
L_P and capacitance C_P are package values, and r_B is the series resistance, which
includes the contact and bulk resistance. The resistance r_d and capacitance C_j are
values defined by equations introduced in earlier sections. For many applications, an
excellent approximate equivalent circuit simply includes an ideal diode in parallel
with the junction capacitance as shown in Fig. 20.4.

A number of hot-carrier rectifiers manufactured by Motorola Semiconductor
Products, Inc., appear in Fig. 20.5 with their specifications and terminal identifica-

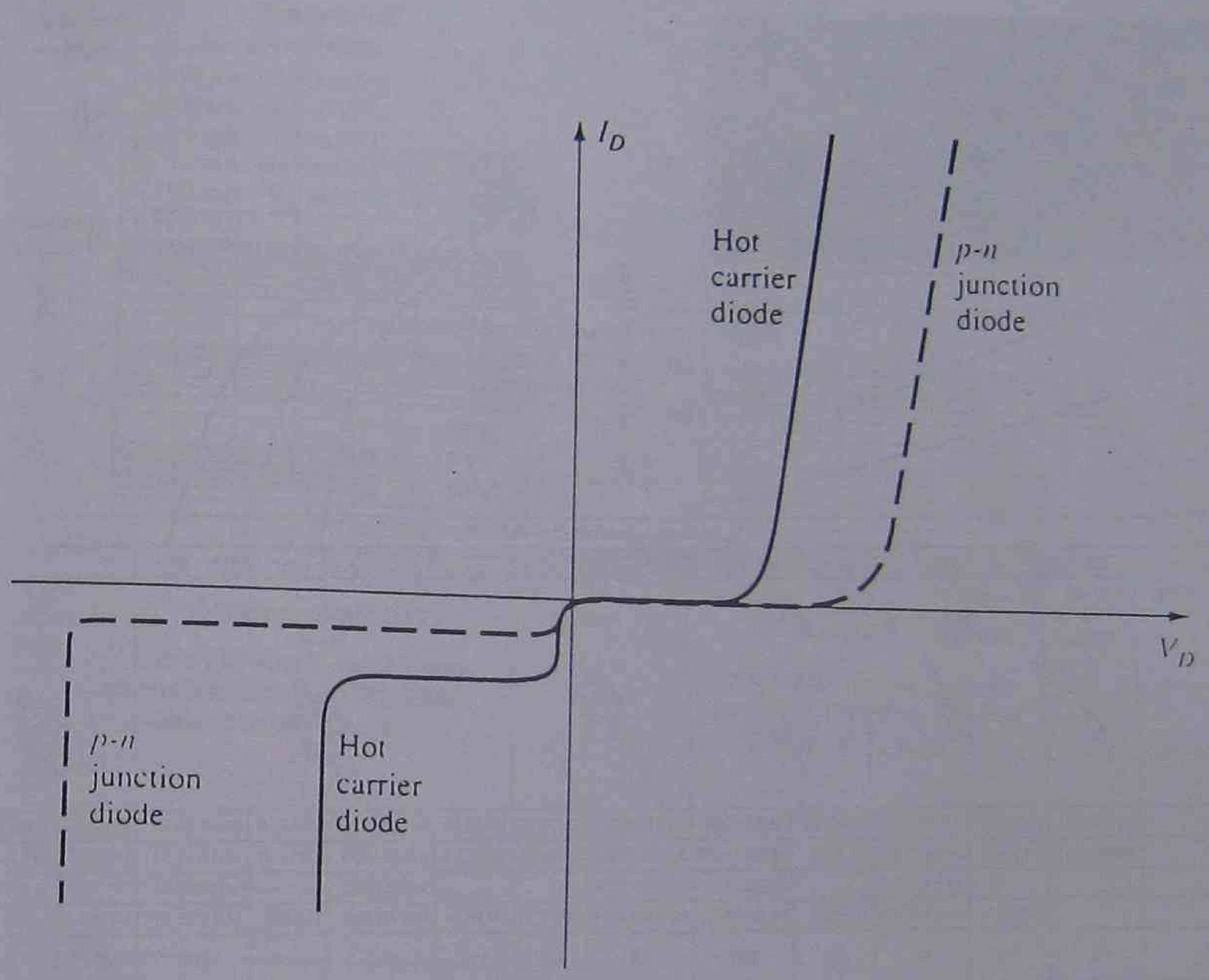


Figure 20.2 Comparison of characteristics of hot-carrier and *p-n* junction diodes.

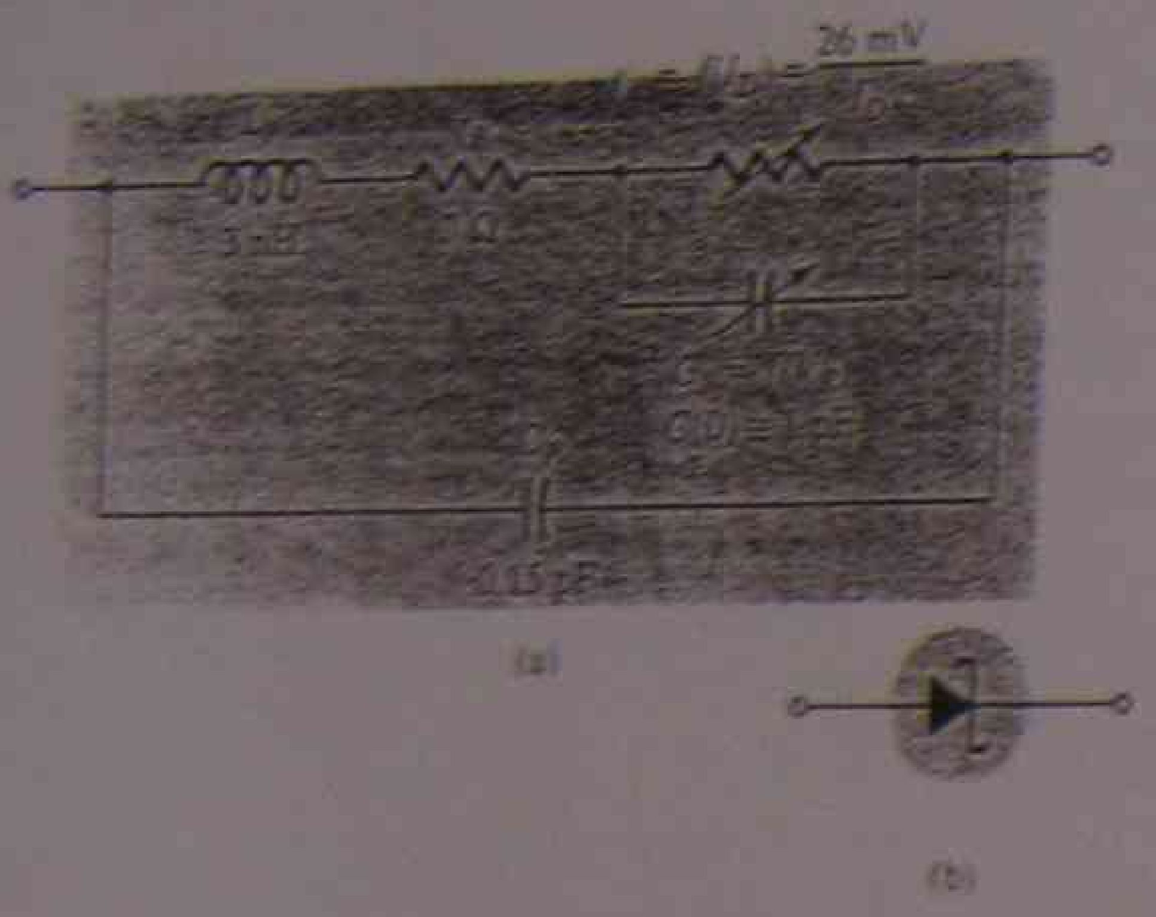


Figure 20.3 Schottky (hot-carrier) diode: (a) equivalent circuit; (b) symbol.

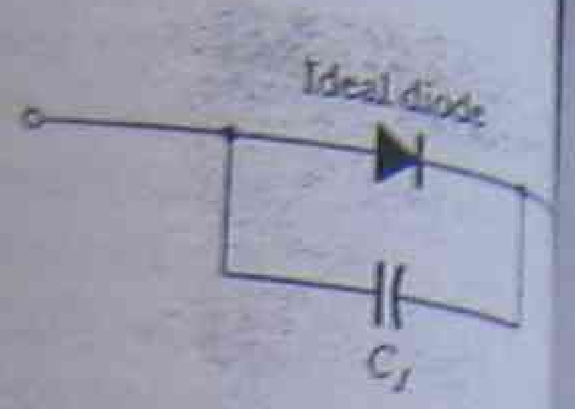
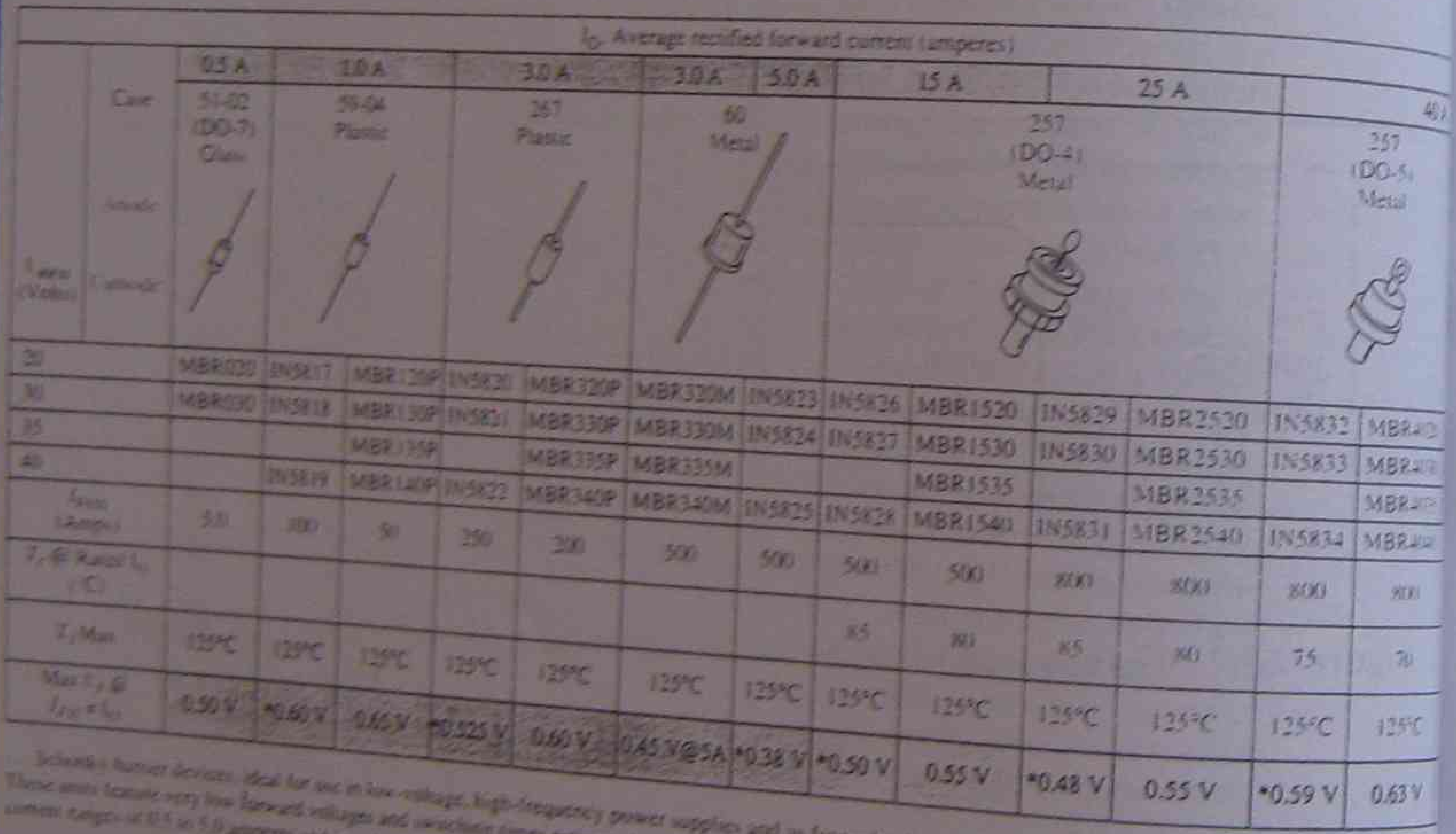


Figure 20.4 Approxima equivalent circuit for the Sc diode



		I_O, Average rectified forward current (amperes)												
		0.5 A	1.0 A		3.0 A		3.0 A	5.0 A	15 A		25 A		40 A	
V_{RRM} (Volts)	Case Anode Cathode	51-02 (DO-7) Glass	59-04 Plastic		267 Plastic		60 Metal		257 (DO-4) Metal				257 (DO-5) Metal	
20		MBR020	1N5817	MBR120P	1N5820	MBR320P	MBR320M	1N5823	1N5826	MBR1520	1N5829	MBR2520	1N5832	MBR40
30		MBR030	1N5818	MBR130P	1N5821	MBR330P	MBR330M	1N5824	1N5827	MBR1530	1N5830	MBR2530	1N5833	MBR40
35				MBR135P		MBR335P	MBR335M			MBR1535		MBR2535		MBR40
40			1N5819	MBR140P	1N5822	MBR340P	MBR340M	1N5825	1N5828	MBR1540	1N5831	MBR2540	1N5834	MBR40
I_{FSM} (Amps)		50	300	50	250	300	500	500	500	500	800	800	800	800
T_C @ Rated I_O (°C)									85	80	85	80	75	70
T_J Max		125°C	125°C	125°C	125°C	125°C	125°C	125°C	125°C	125°C	125°C	125°C	125°C	125°C
Max V_F @ $I_{FM} = I_O$		0.50 V	*0.60 V	0.65 V	*0.525 V	0.60 V	0.45 V@5A	*0.38 V	*0.50 V	0.55 V	*0.48 V	0.55 V	*0.59 V	0.63 V

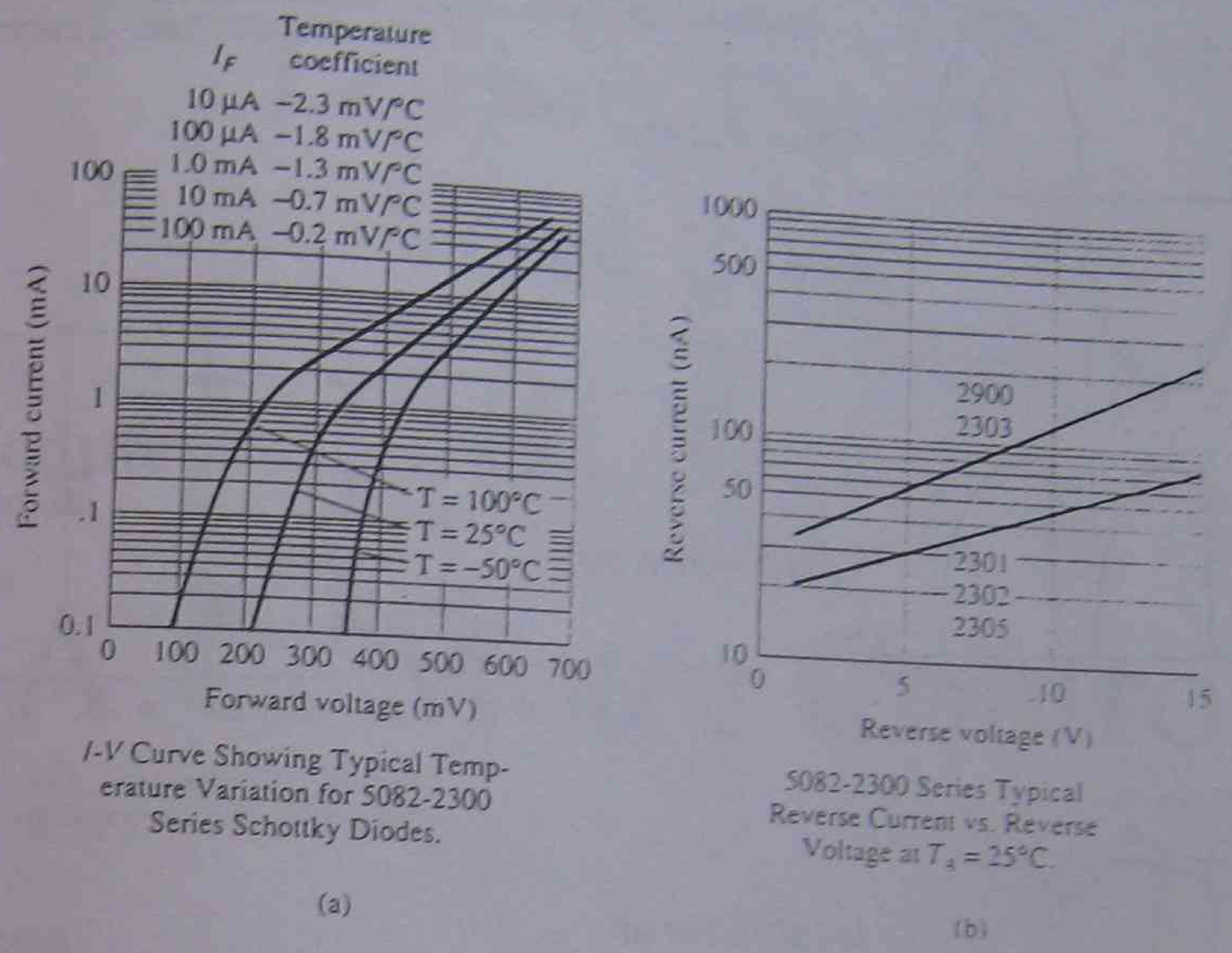
Schottky barrier devices, ideal for use in low-voltage, high-frequency power supplies and as free-wheeling diodes. These units feature very low forward voltages and switching times estimated at less than 10 ns. They are offered in current ranges of 0.5 to 5.0 amperes and in voltages to 40.

V_{RRM} =repetitive peak reverse voltage
I_{FSM} =forward current, surge peak
$I_{F(AV)}$ =forward current, maximum

Figure 20.5 Motorola Schottky barrier devices. (Courtesy Motorola Semiconductor Products, Incorporated.)

tion. Note that the maximum forward voltage drop V_F does not exceed 0.65 V for any of the devices, while this was essentially V_T for a silicon diode.

Three sets of curves for the Hewlett-Packard 5082-2300 series of general-purpose Schottky barrier diodes are provided in Fig. 20.6. Note at $T = 100°C$ in Fig. 20.6a that V_F is only 0.1 V at a current of 0.01 mA. Note also that the reverse current has been limited to nanoamperes in Fig. 20.6b and the capacitance to 1 pF in Fig. 20.6c to ensure a high switching rate.

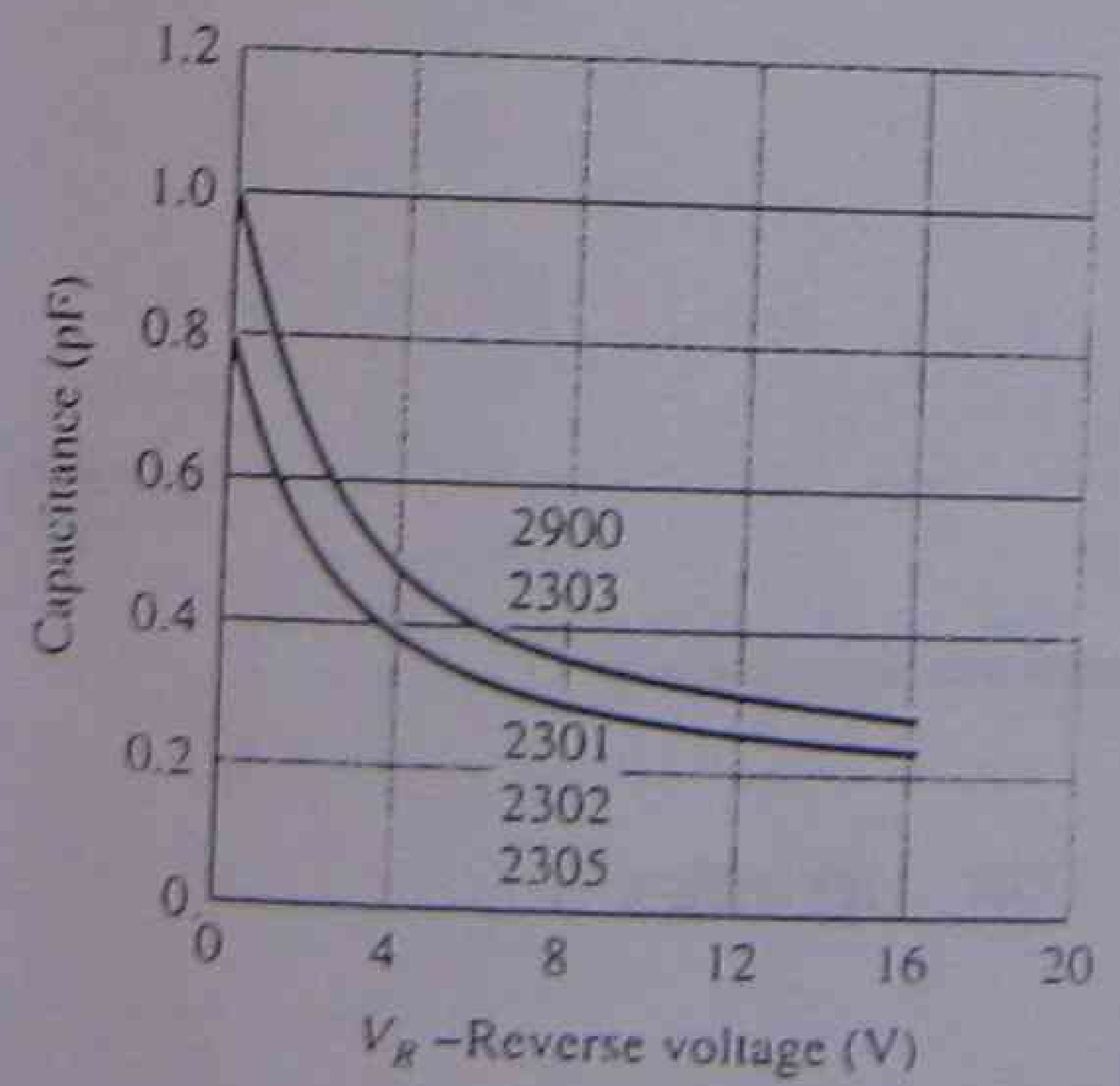


I-V Curve Showing Typical Temperature Variation for 5082-2300 Series Schottky Diodes.

(a)

5082-2300 Series Typical Reverse Current vs. Reverse Voltage at $T_A = 25°C$.

(b)



5082-2300 Series Typical Capacitance vs. Reverse Voltage at $T_A = 25°C$.

(c)

Figure 20.6 Characteristic curves for Hewlett-Packard 5082-2300 series of general-purpose Schottky barrier diodes. (Courtesy Hewlett-Packard Corporation.)

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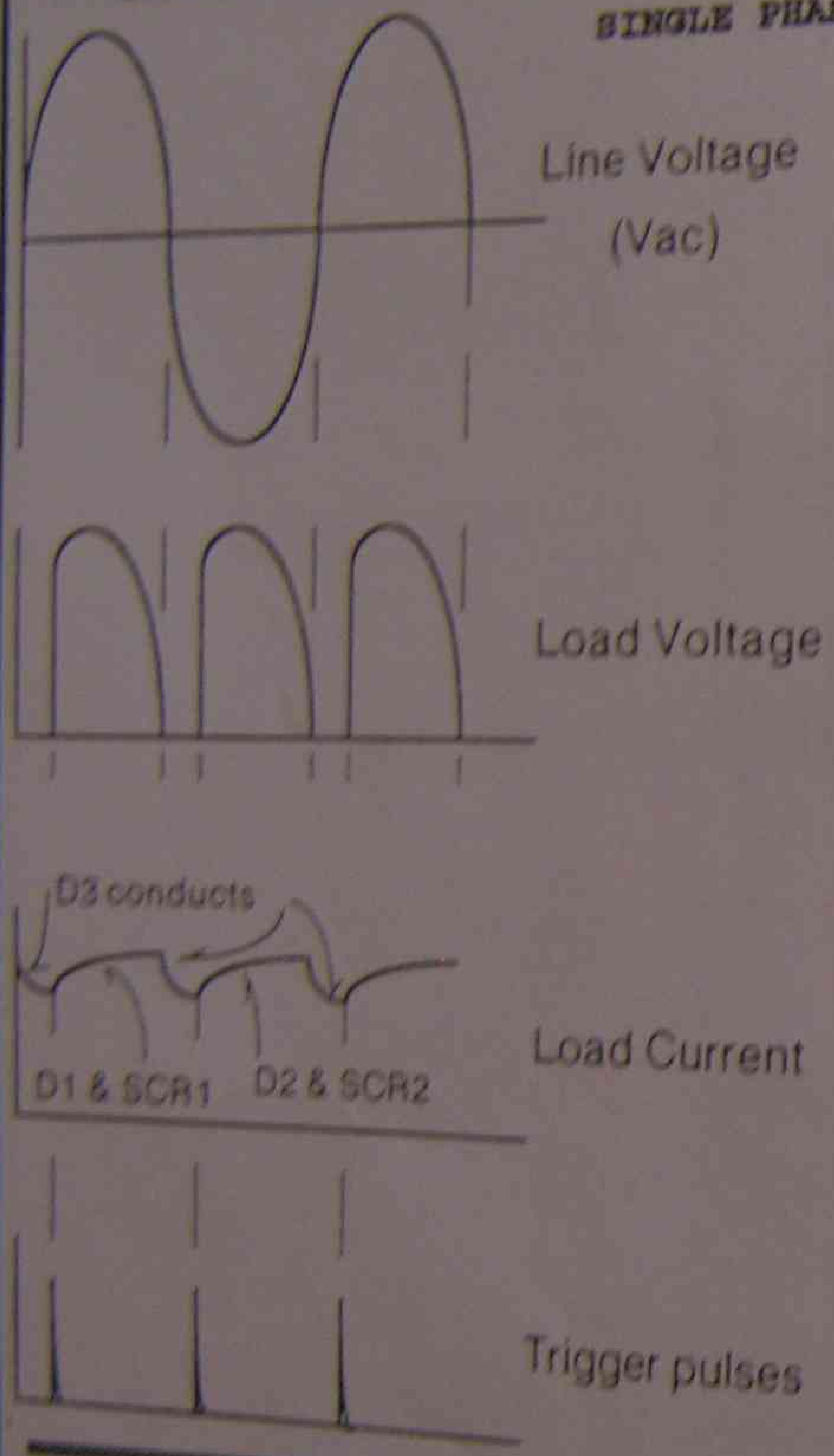
SINGLE PHASE, HALF CONTROLLED, BRIDGE RECTIFIER CIRCUI



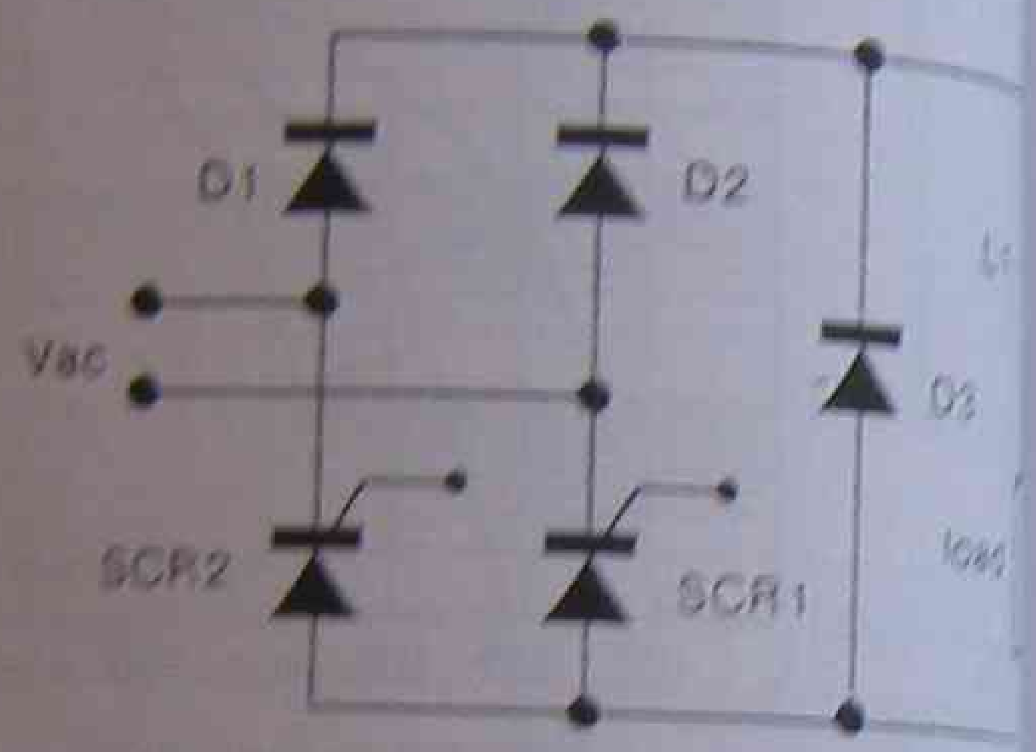


Fig.10 Single-phase, half co bridge with flywheel diode (D3)

SUITABLE FOR SEVERAL kW SUPPLIES.

USE IN A 3 PHASE CONFIGURATION FOR HIGHER POWERS.

EXHIBITS A HIGH POWER FACTOR.

FOR INDUCTIVE LOADS D3, FREE WHEELING DIODE, MUST BE USED.

ALSO FOR INDUCTIVE LOADS TO MINAMISE di/dt THE SATURABLE REACTOR, L1, IS INCORPORATED.

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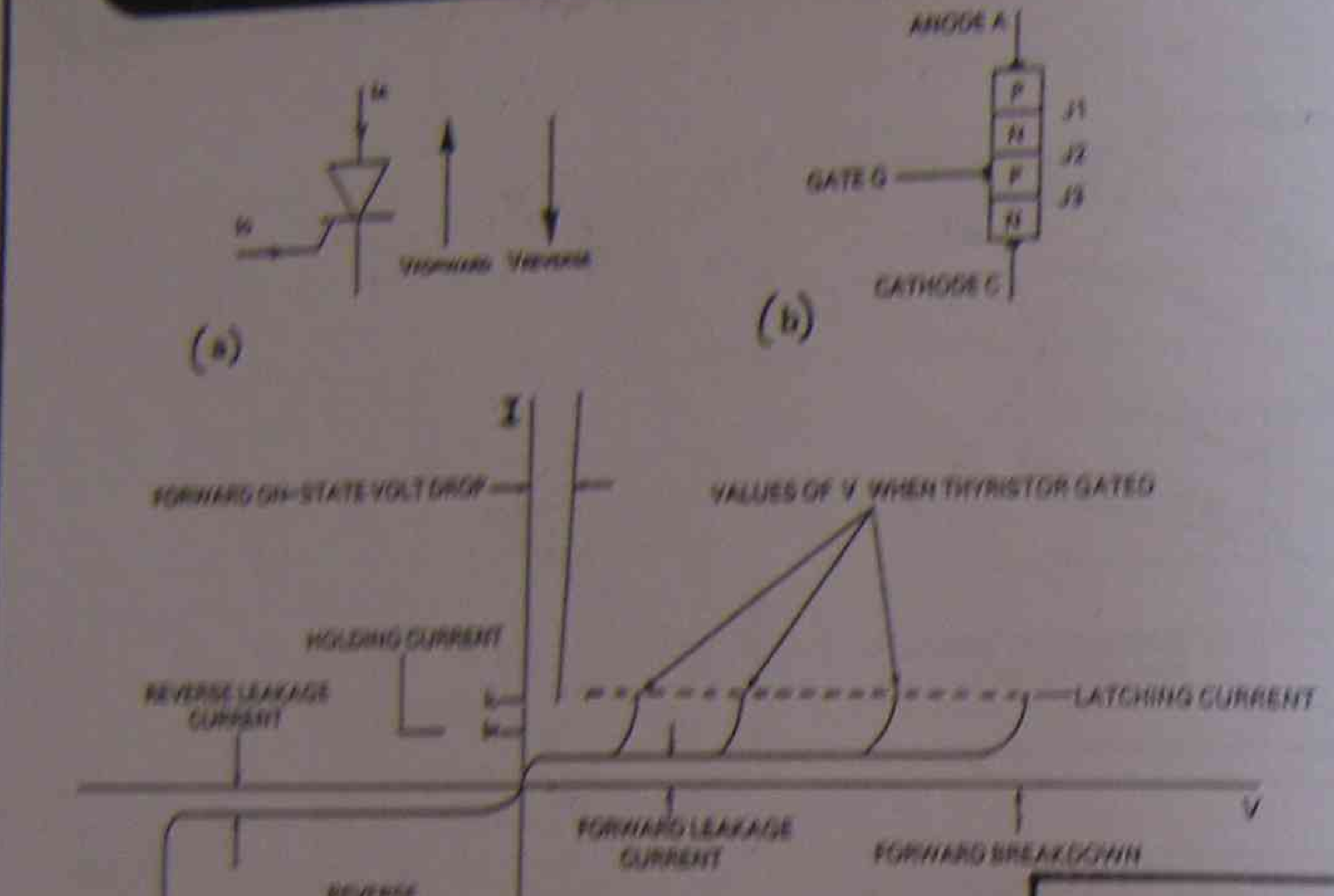
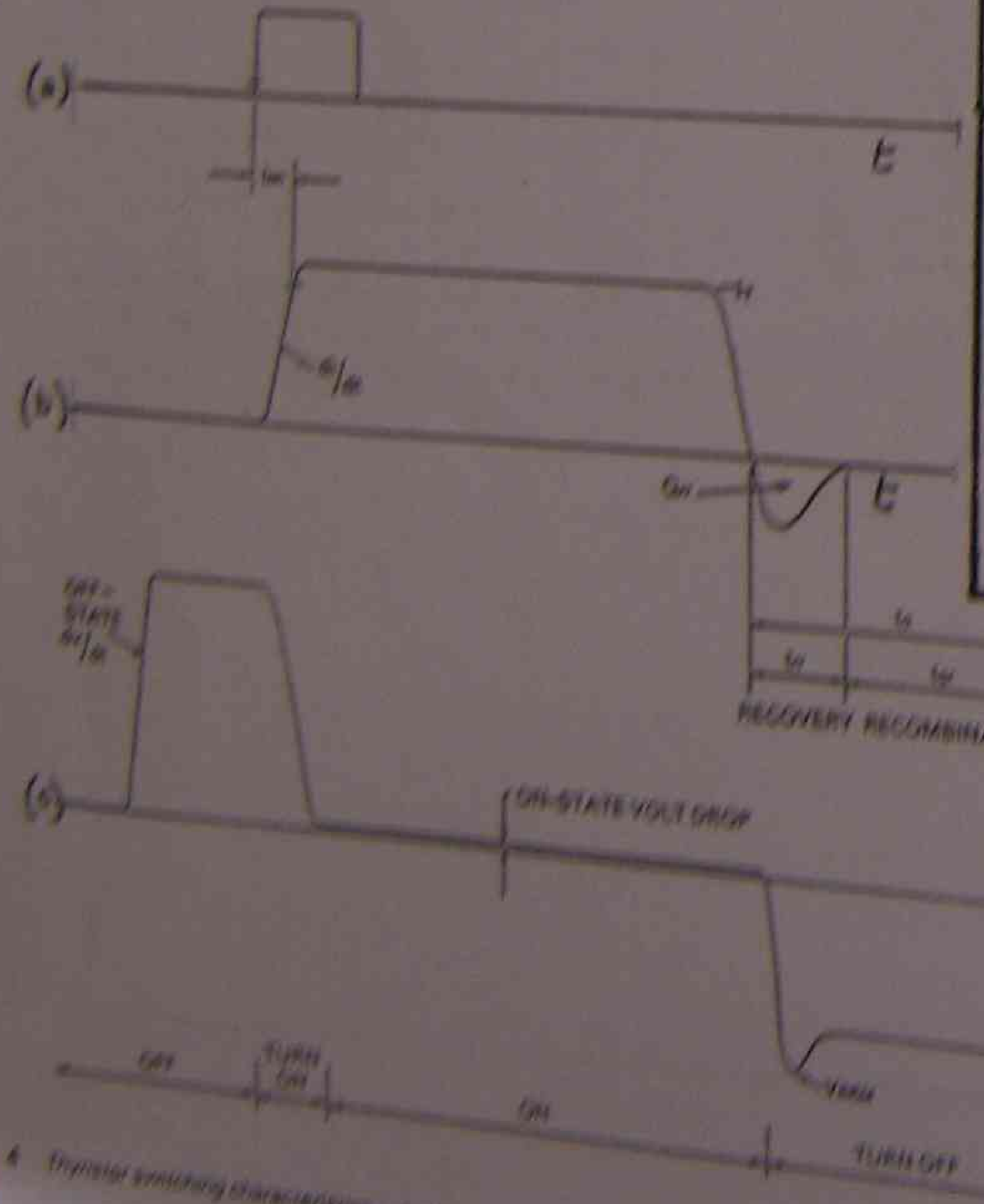
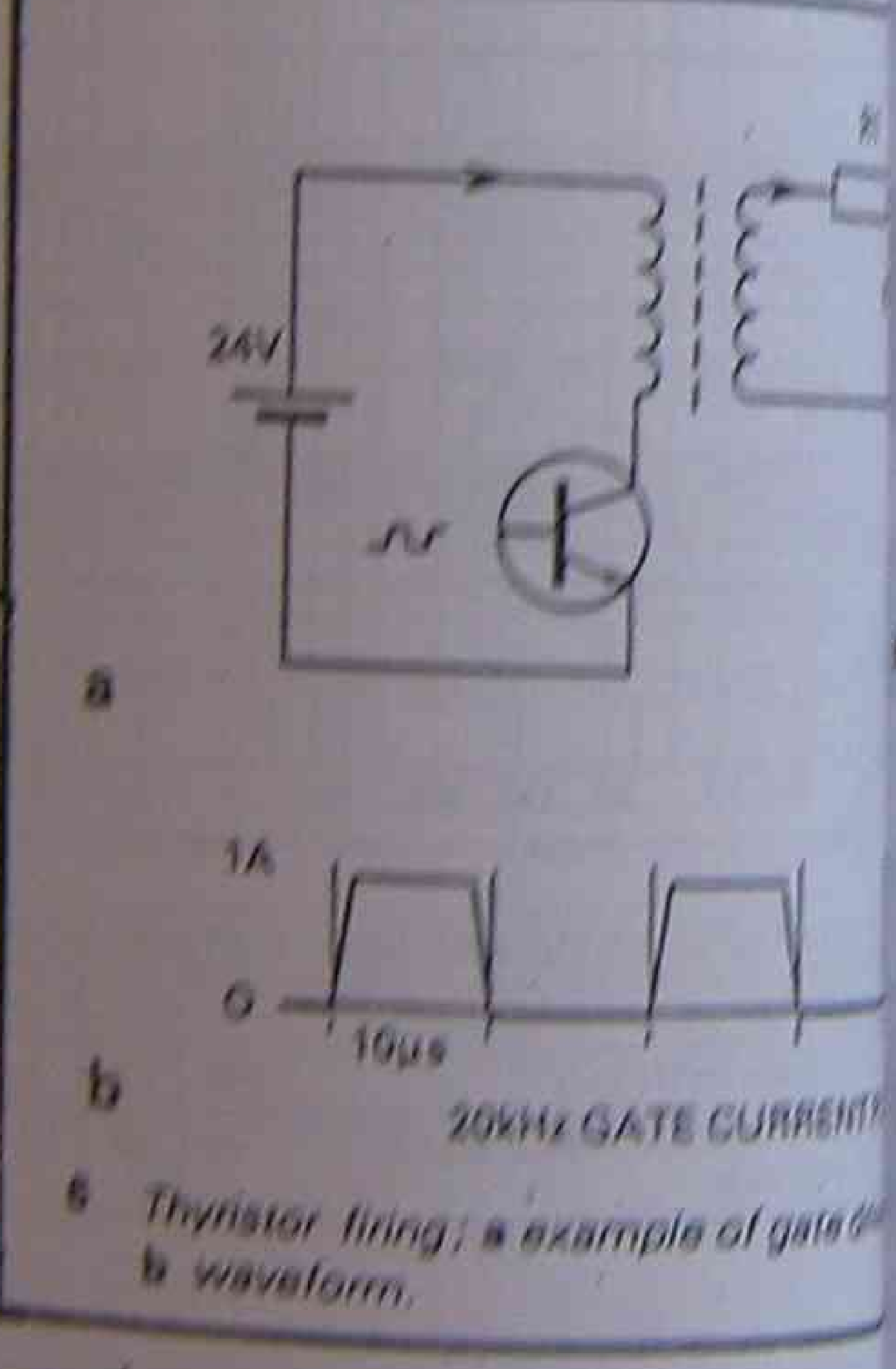


3 Thyristor: a circuit symbol; b structure; c characteristic.

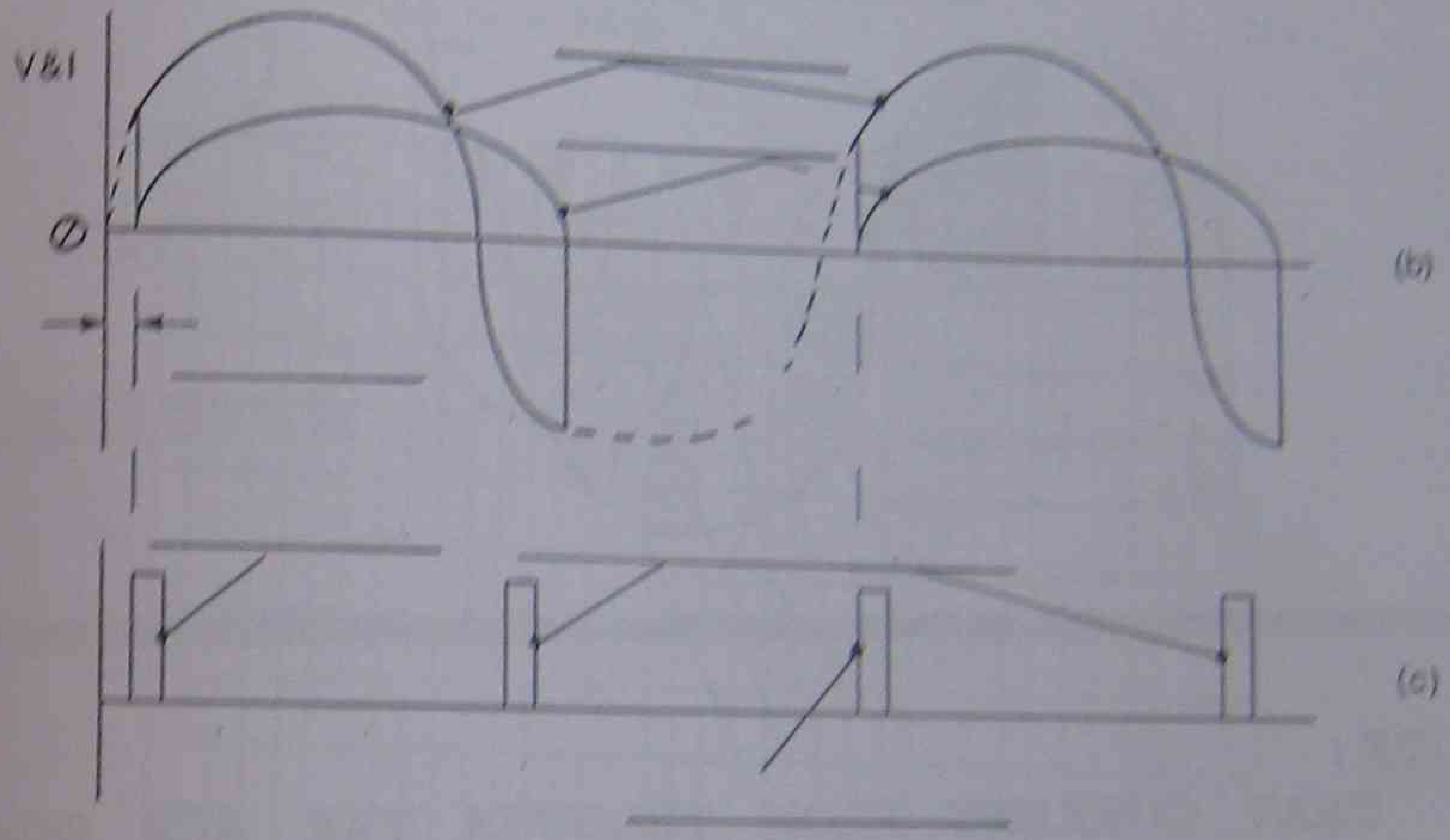


4 Thyristor switching characteristics : a gate current; b anode current; c anode voltage



5 Thyristor firing; a example of gate d
b waveform.

FIG 3(b) & (c),
VOLTAGE AND CURRENT WAVEFORMS FOR AN INDUCTIVE LOAD WITH SMALL CONDUCTION ANGLE.



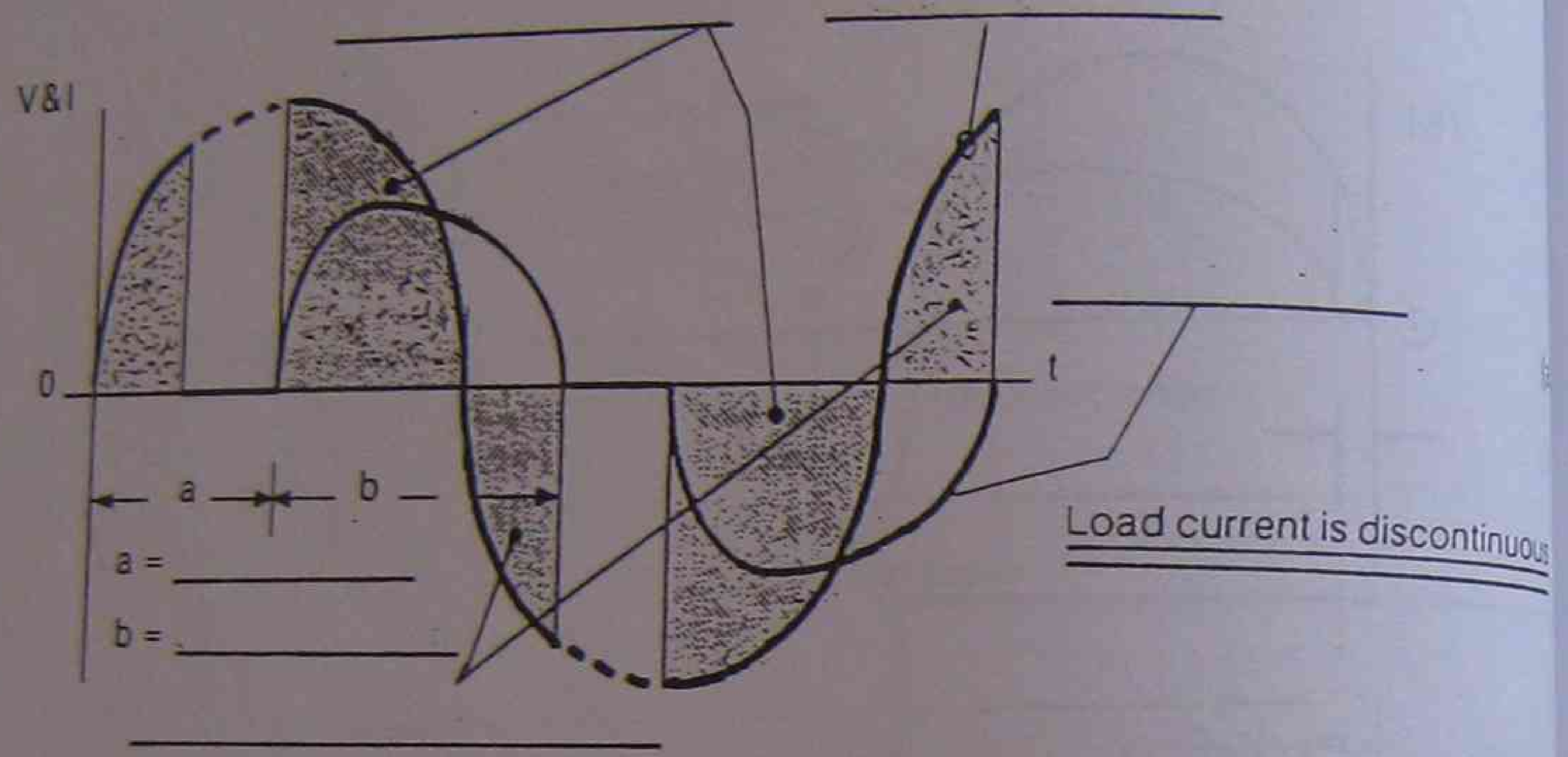
THE SITUATION SHOWN IN (a) LEADS TO PROBLEMS WHEN THE FIRING ANGLE IS SMALL.

AFTER INITIAL FIRING OF SCR1 THE CONDUCTION OF SCR2 IS PREVENTED BECAUSE OF SCR1 STILL CONDUCTING AND REVERSE BIASING IT.

THIS SITUATION IS THE CASE WITH A SIMPLE SINGLE PULSE FIRING SYSTEM.

WITH THE USE OF AN EXTENDED PULSE OR A PULSE TRAIN OF SUITABLE DURATION, SCR WILL FIRE AS SOON AS CONDITIONS ALLOW.

FIG 3(a),
VOLTAGE AND CURRENT WAVEFORMS FOR AN INDUC
LOAD.



NOTE:

- THAT CURRENT FLOW THROUGH THE SCR DOES STOP UNTIL THE END OF PERIOD b, THERFOR NATURAL COMMUTATION IS BEING USED S WOULD NOT TURN OFF UNTIL THIS POINT.
- THERE IS TWO WAY FLOW OF ENERGY, BOTH F THE SUPPLY AND BEING RETURNED TO SUPPLY.

LOAD CURVES OF RMS SUPPLY TO RMS LOAD VOLTAGE FOR LOAD POWER FACTORS OF 1.0 (PURE RESISTIVE) TO 0 (PURE INDUCTIVE).

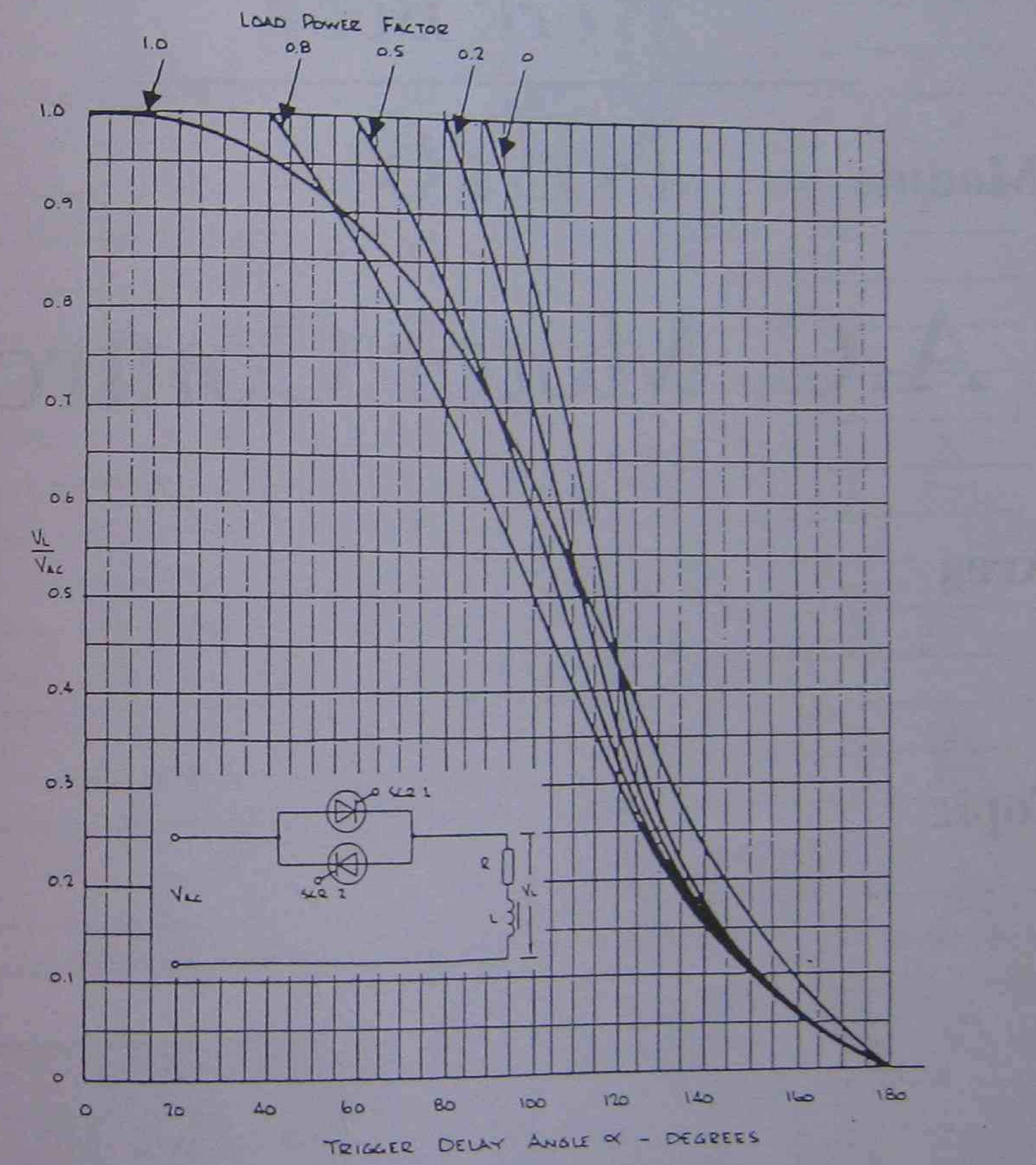


Fig.2 Curves showing the ratio of load to supply voltage for various load power factors and trigger angles

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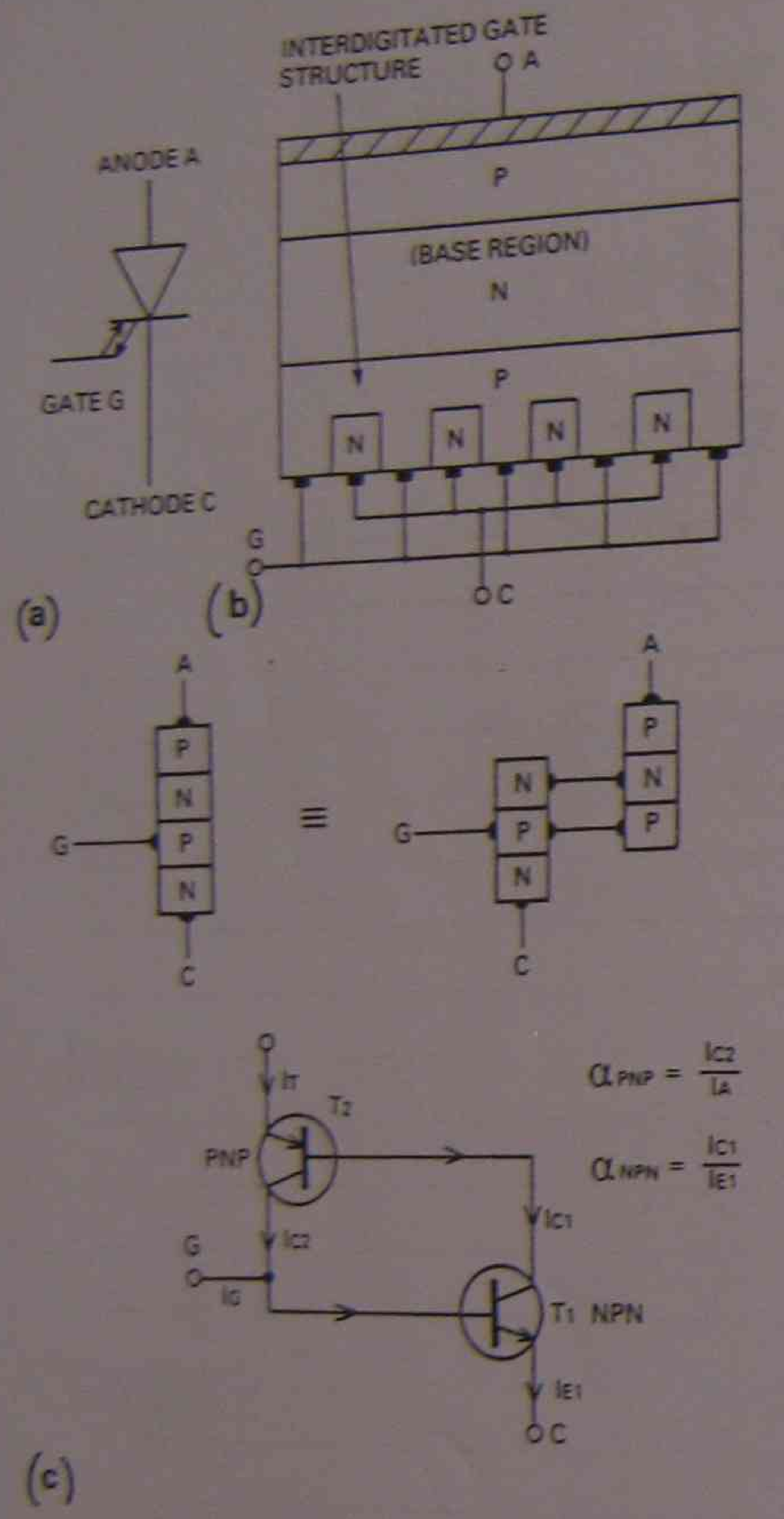


8 *GTO:* **a** *circuit symbol;* **b** *structure;* **c** *two-transistor model of thyristor.*

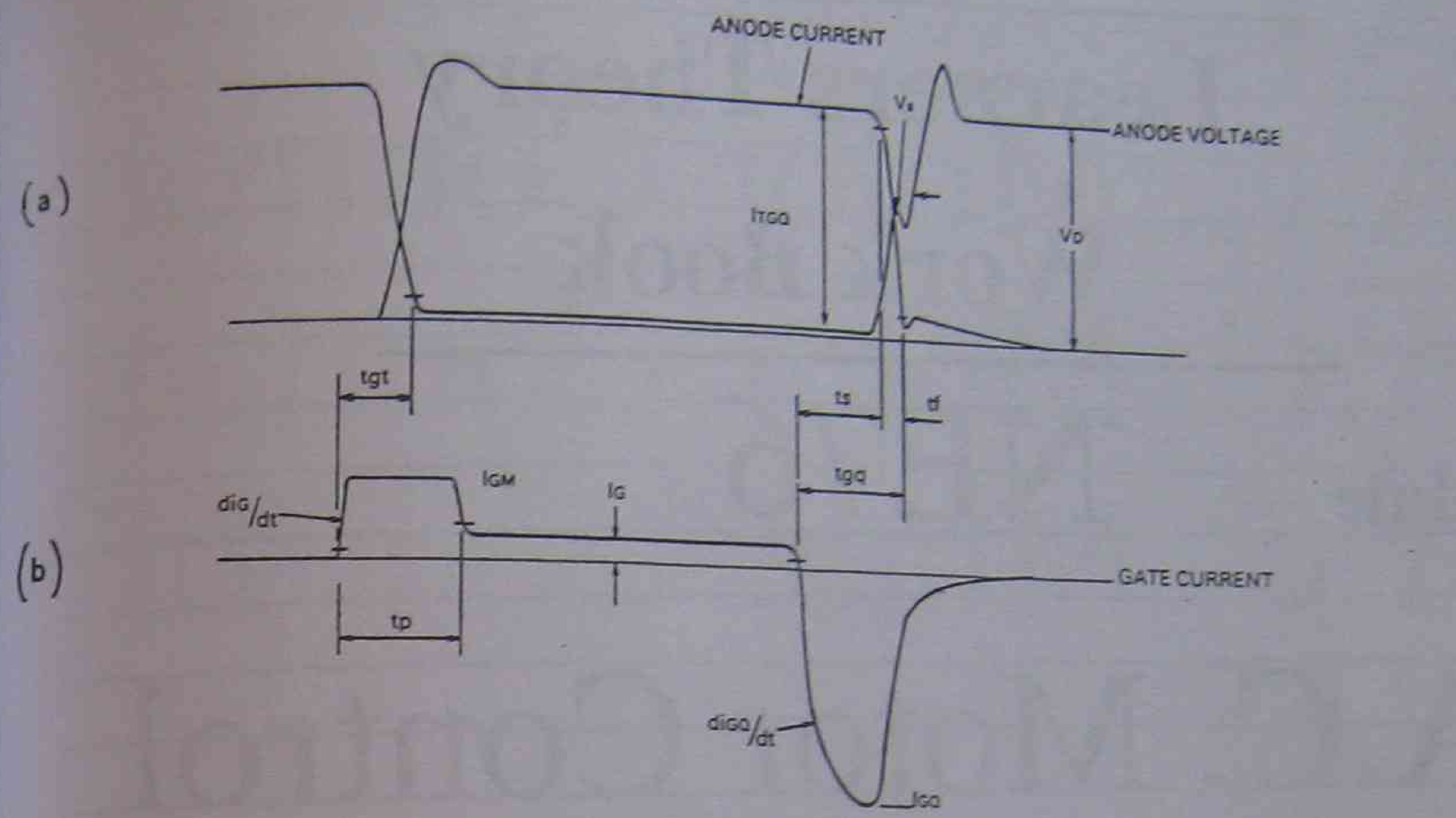
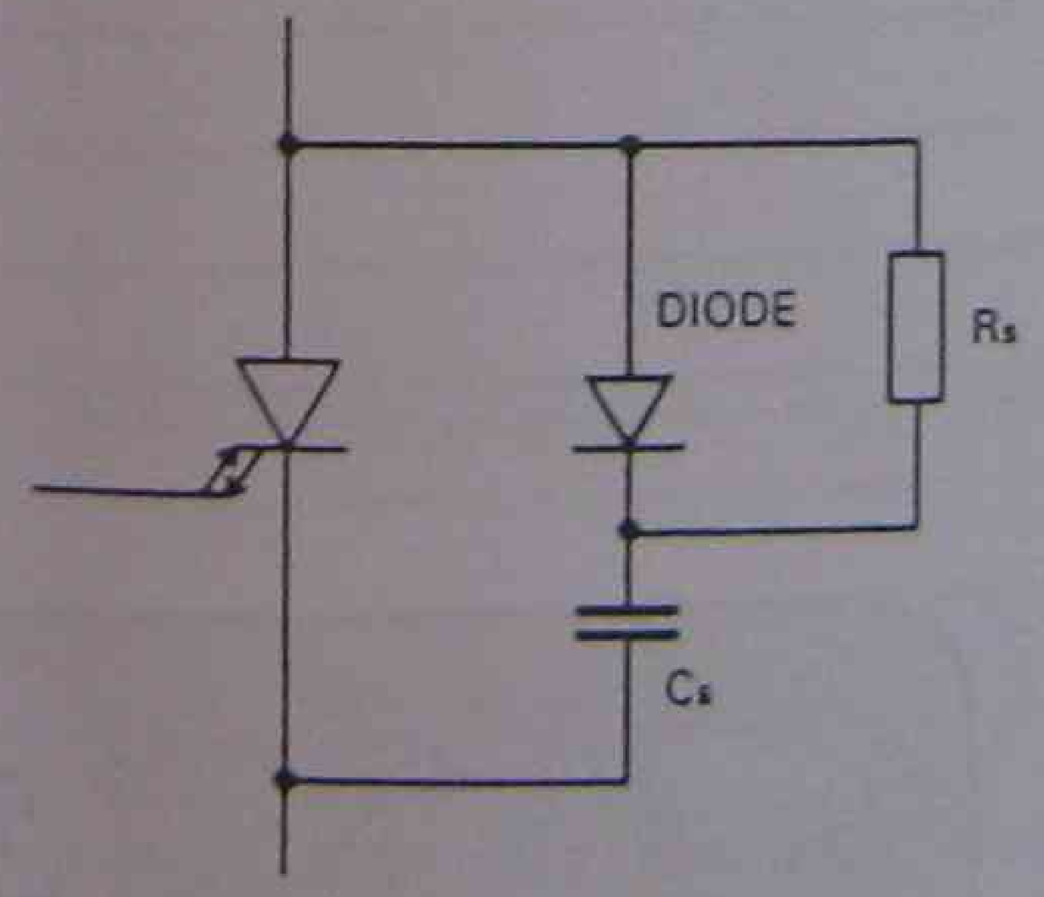


TABLE 1 Recommended Gate Drive for IR 160 PFT

I_{GM} ≥ ,10A
I_G ≥ 2A
I_{GQ} ≥ 120A
tw ≥ 10μs
dI_G/dt = 5A/μs

RESISTIVE SWITCHING PERFORMANCE
V_D = 600V I_{TGQ} = 600A max.
Snubber = 2μF
t_{gt} = 5μs (turn-on time)
t_{gq} = 8μs (turn-off time)
t_f = 0.8μs

9 *GTO switching performance;* **a** *anode current and voltage;* **b** *gate current.*



10 *Typical snubber circuit required for GTO.*

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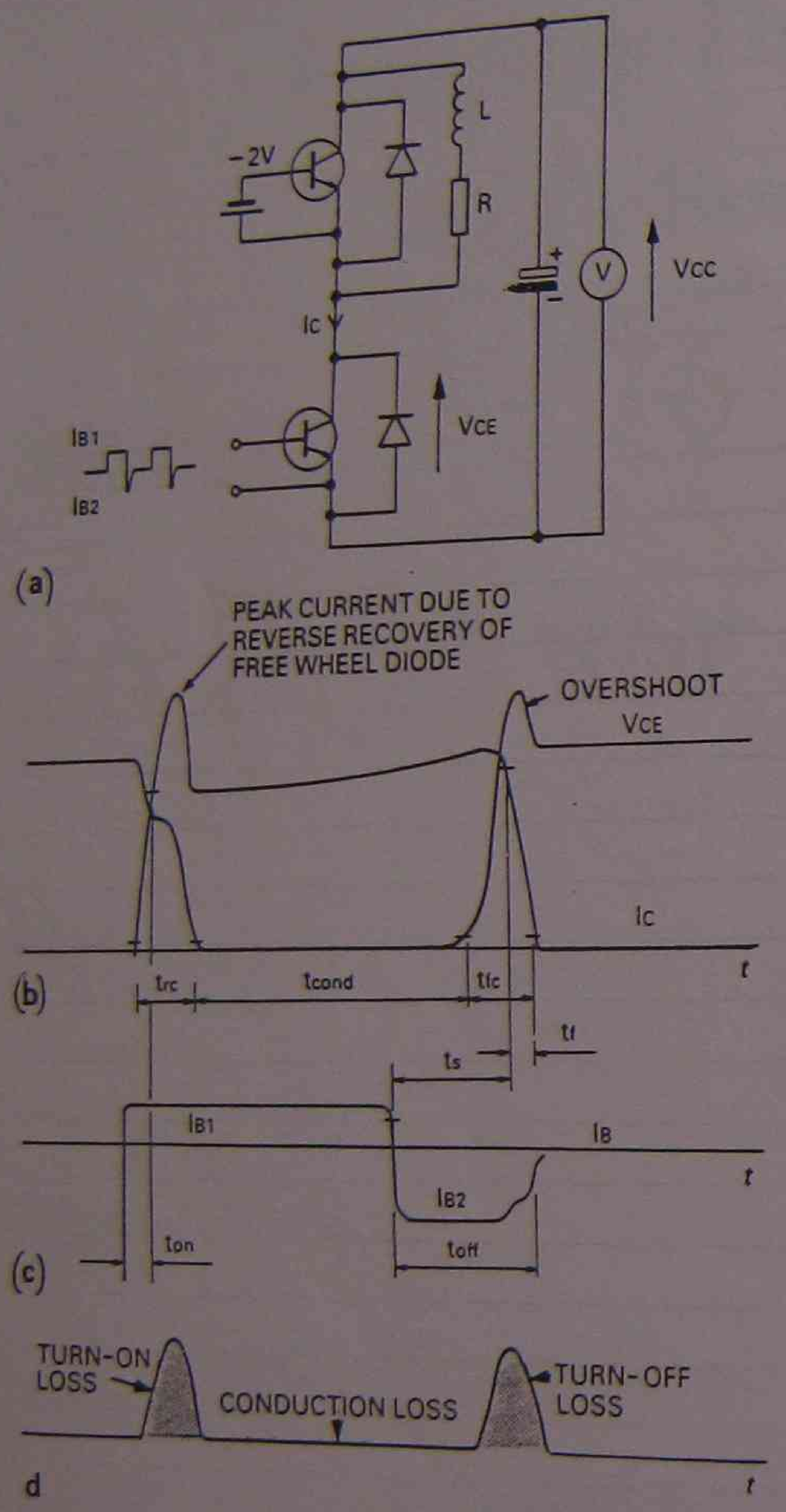
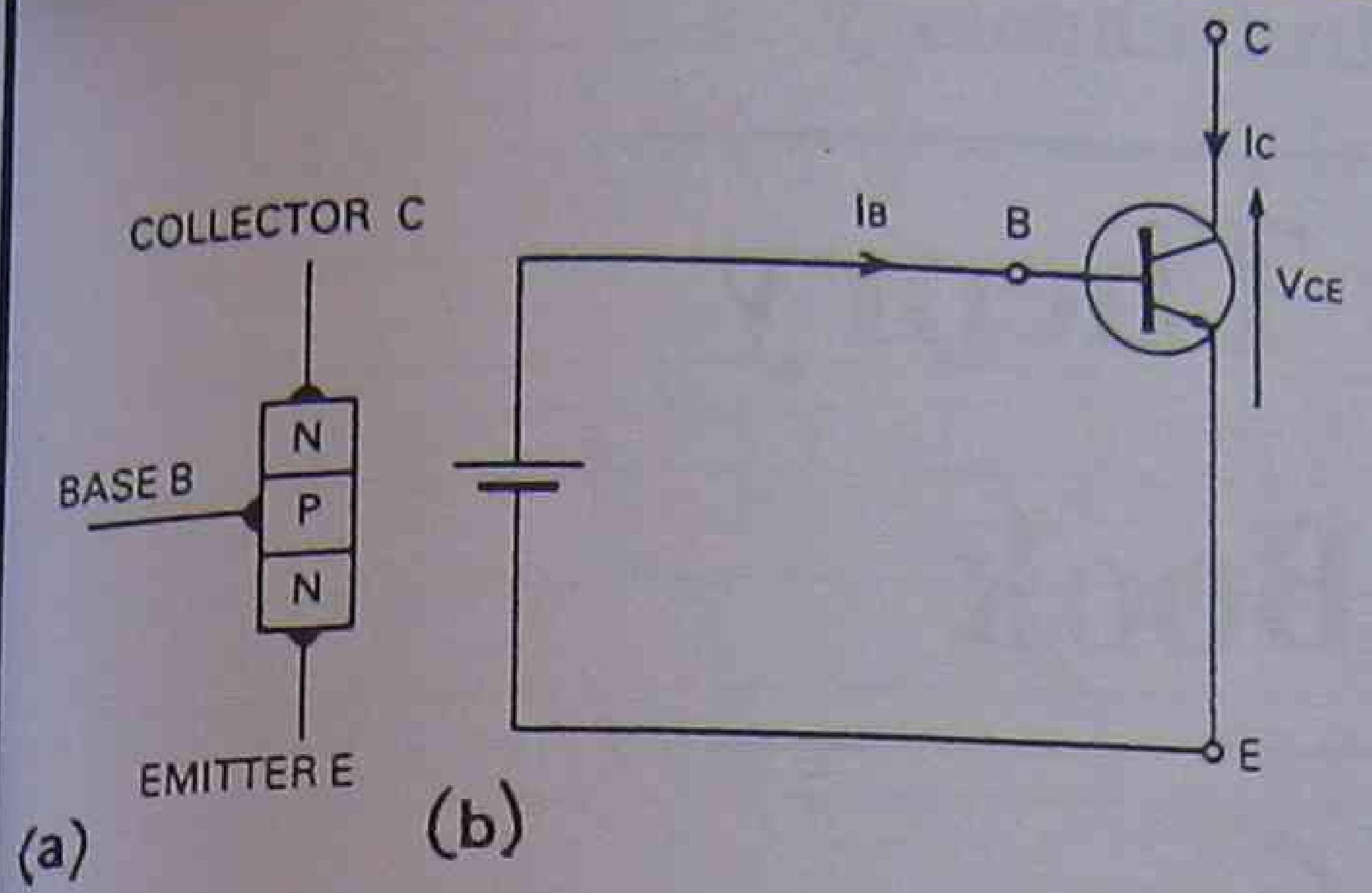
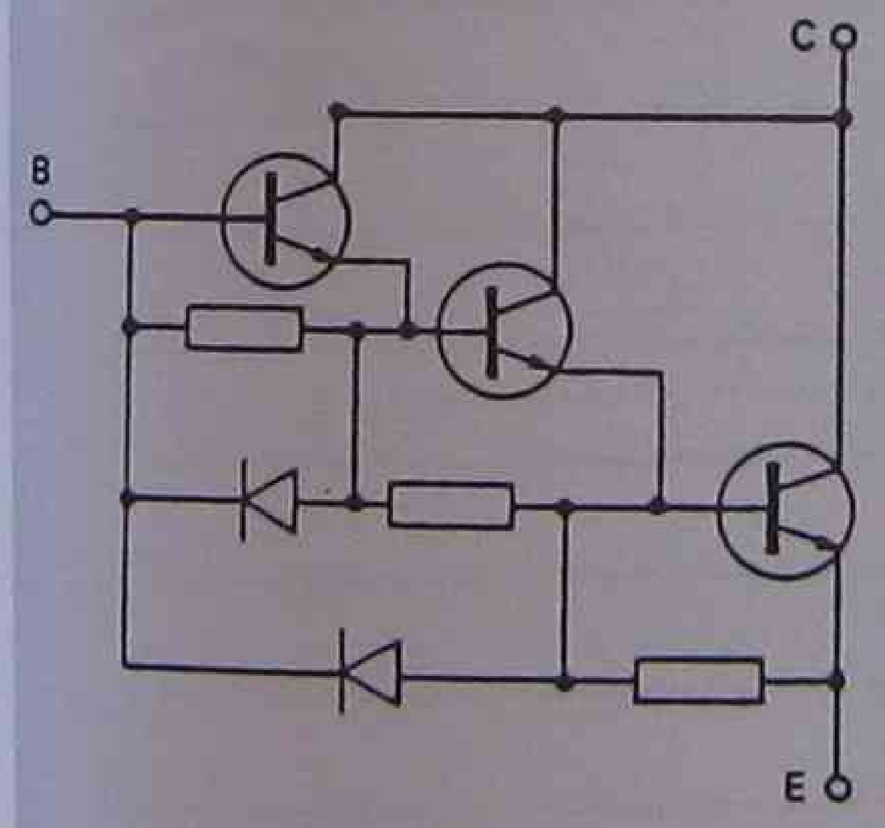


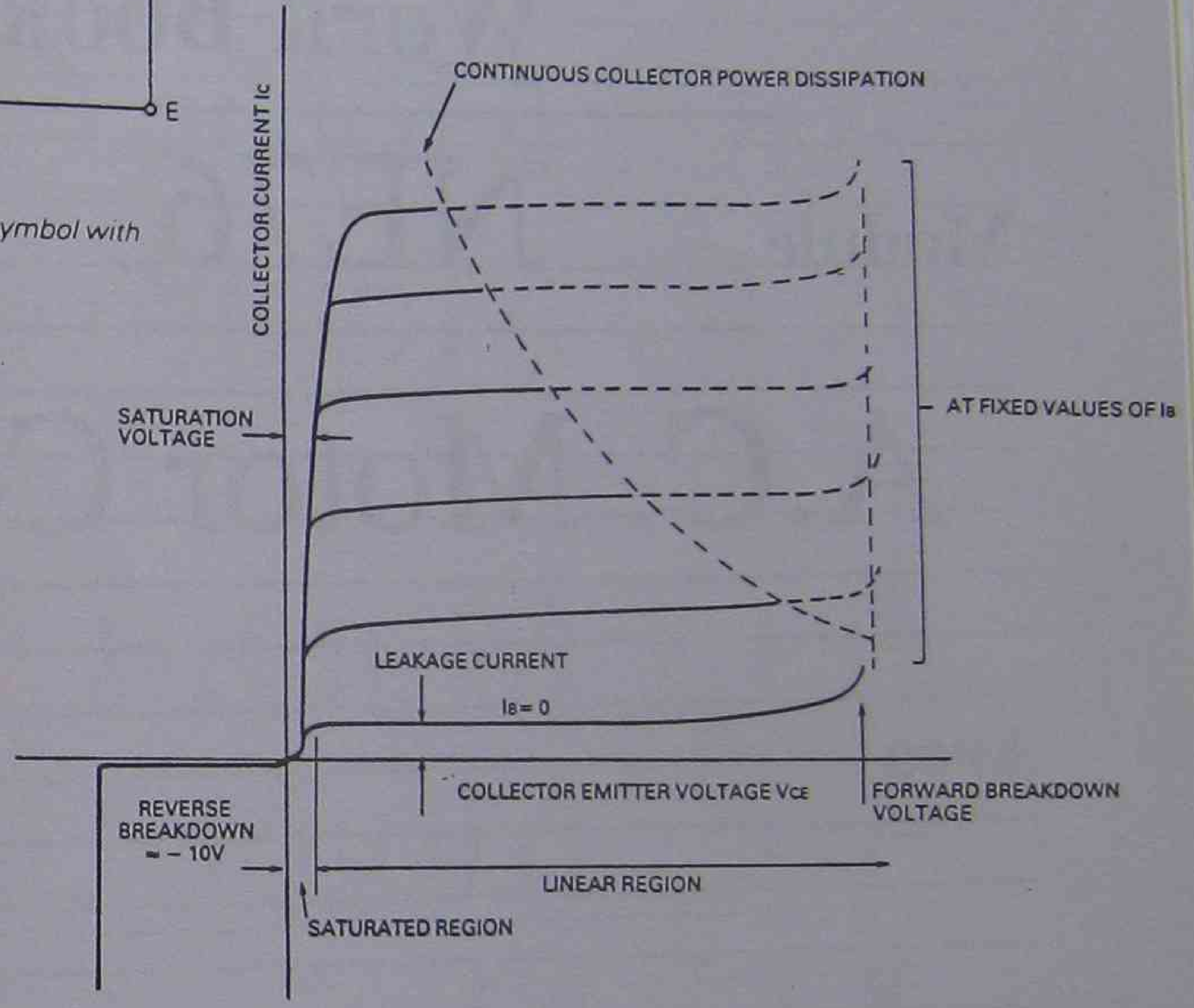
15 *Bipolar transistor switching an inductive load:* **a** *test circuit;* **b** *collector current and voltage waveforms;* **c** *base drive;* **d** *power loss over one switching cycle.*



11 *NPN transistor;* **a** *structure;* **b** *circuit symbol with current directions.*



13 *Three-stage Darlington arrangement of power transistor.*



12 *Output characteristic (common emitter) for NPN transistor*

TABLE 2 Calculation of Switching and Conduction Losses for a Bipolar Transistor

Energy dissipated during turn-on	$W_{on} \approx \frac{1}{2} t_{rc} \times V_{CC} \times I_C$
Energy dissipated during turn-off	$W_{off} \approx \frac{1}{2} t_{fc} \times V_{CC} \times I_C$
Energy dissipated during conduction period	$W_{cond} \approx V_{CEsat} \times I_C \times t_{cond}$
Switching loss (watts)	$f(W_{on} + W_{off})$
Conduction loss (watts)	$f \times W_{cond}$
TOTAL AVERAGE POWER LOSS	Switching loss + Conduction loss

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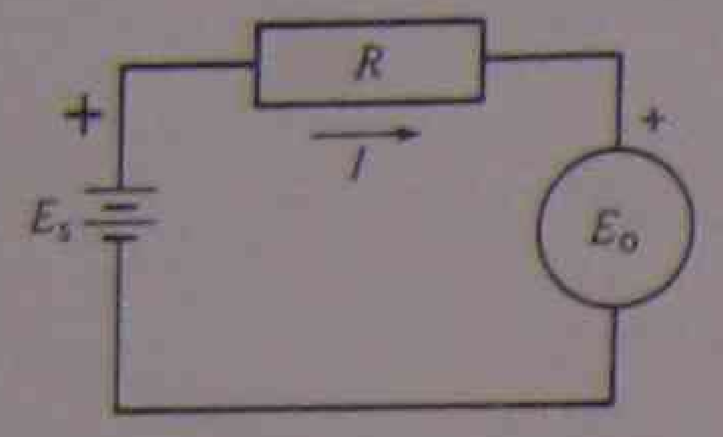


Figure 21-38
Inefficient power transfer.

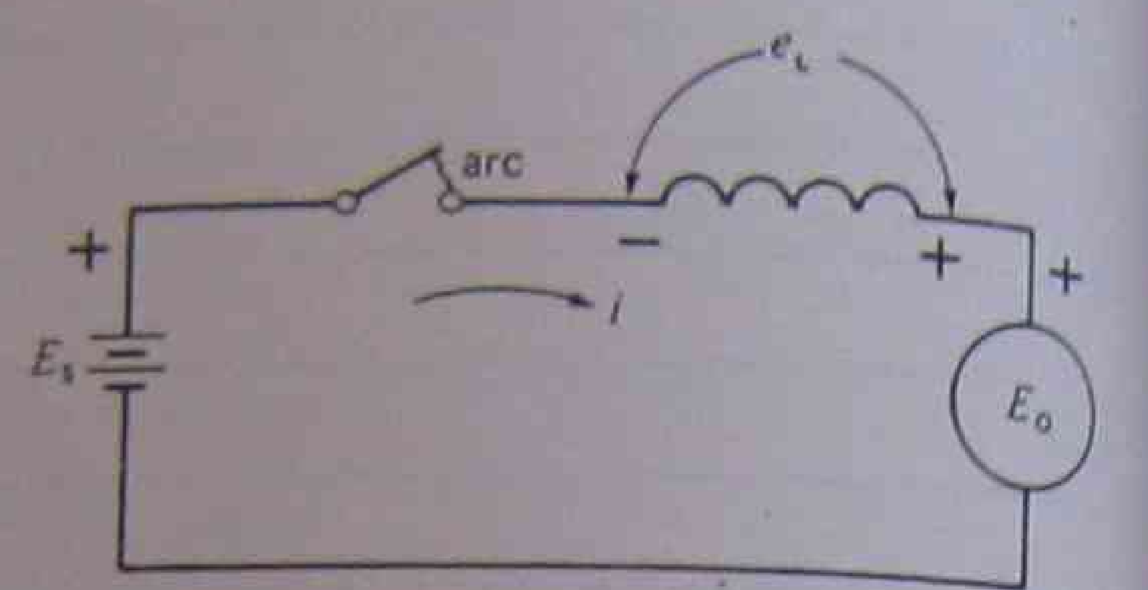


Figure 21-41
Energy is dissipated in the arc. Note the polarity of e_L.

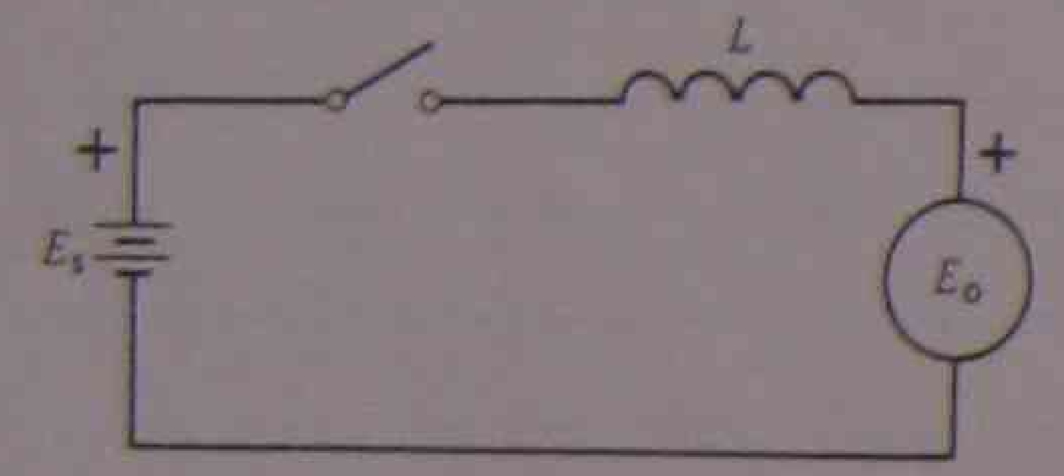


Figure 21-39
Energy transfer using an inductor.

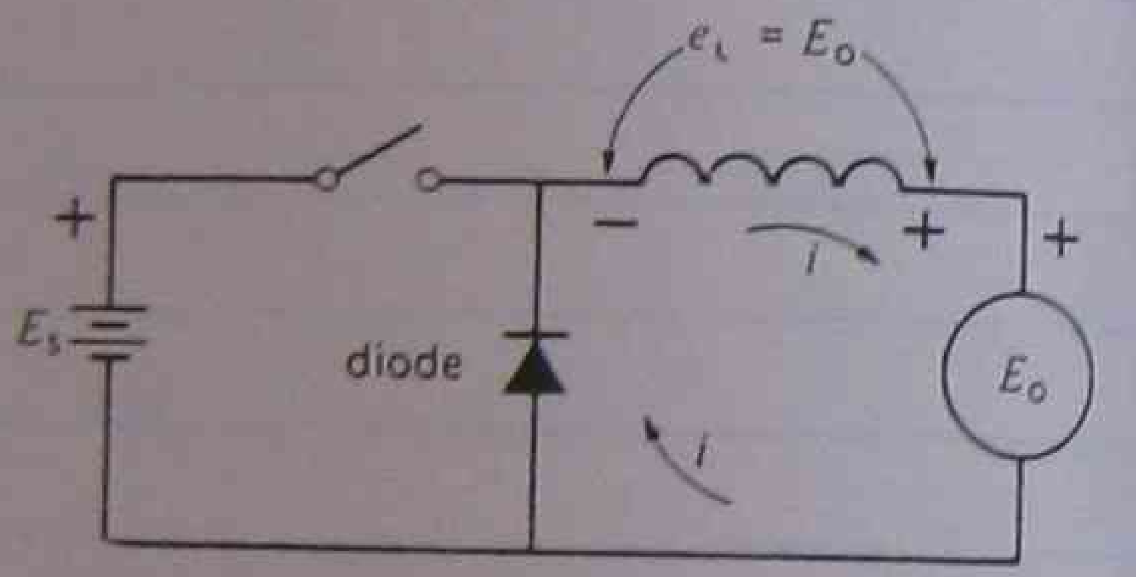


Figure 21-42
Energy transferred without loss.

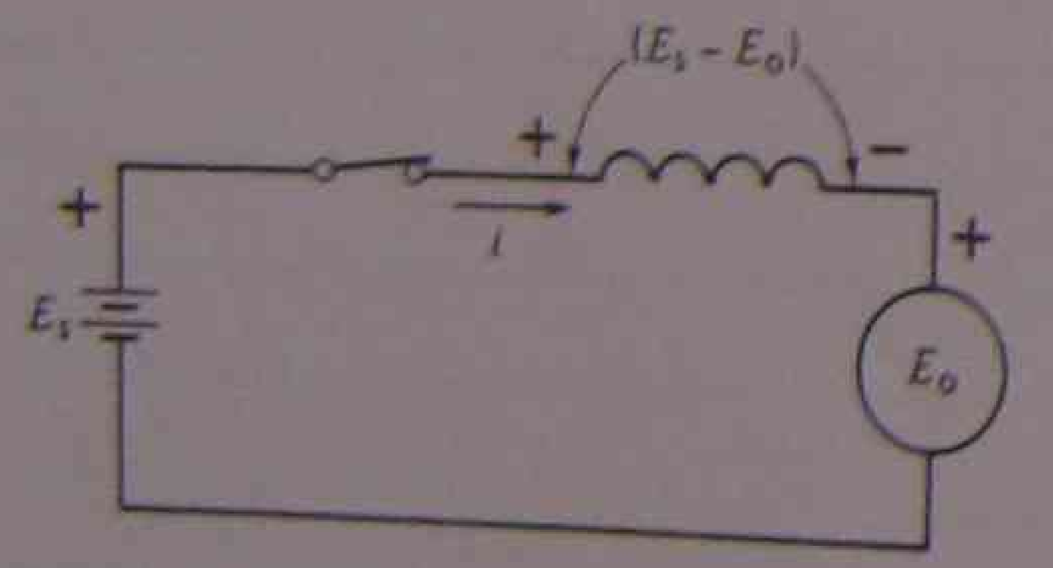


Figure 21-40
Energy is stored in the inductor.

$(A+)T_1 = (A-)T_2$

$(E_s - E_o)T_1 = E_0T_2$

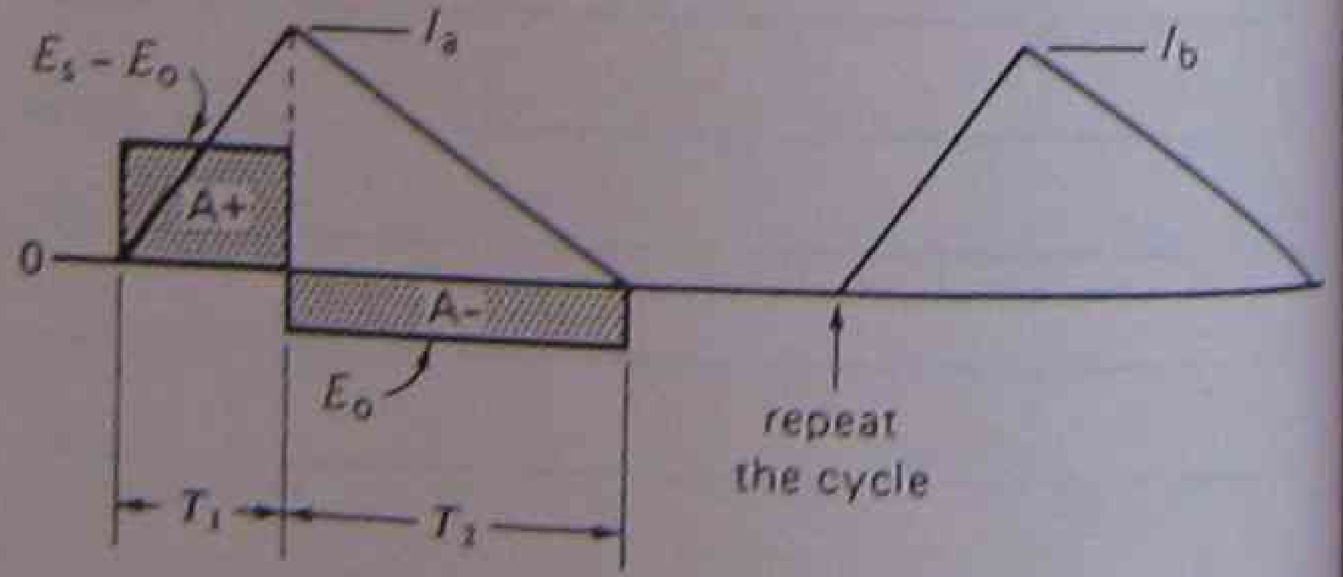


Figure 21-43
E and I in the inductor of Fig. 21-42.

at T_1 $I_a = A+/L$

$= (E_s - E_0)T_1/L$

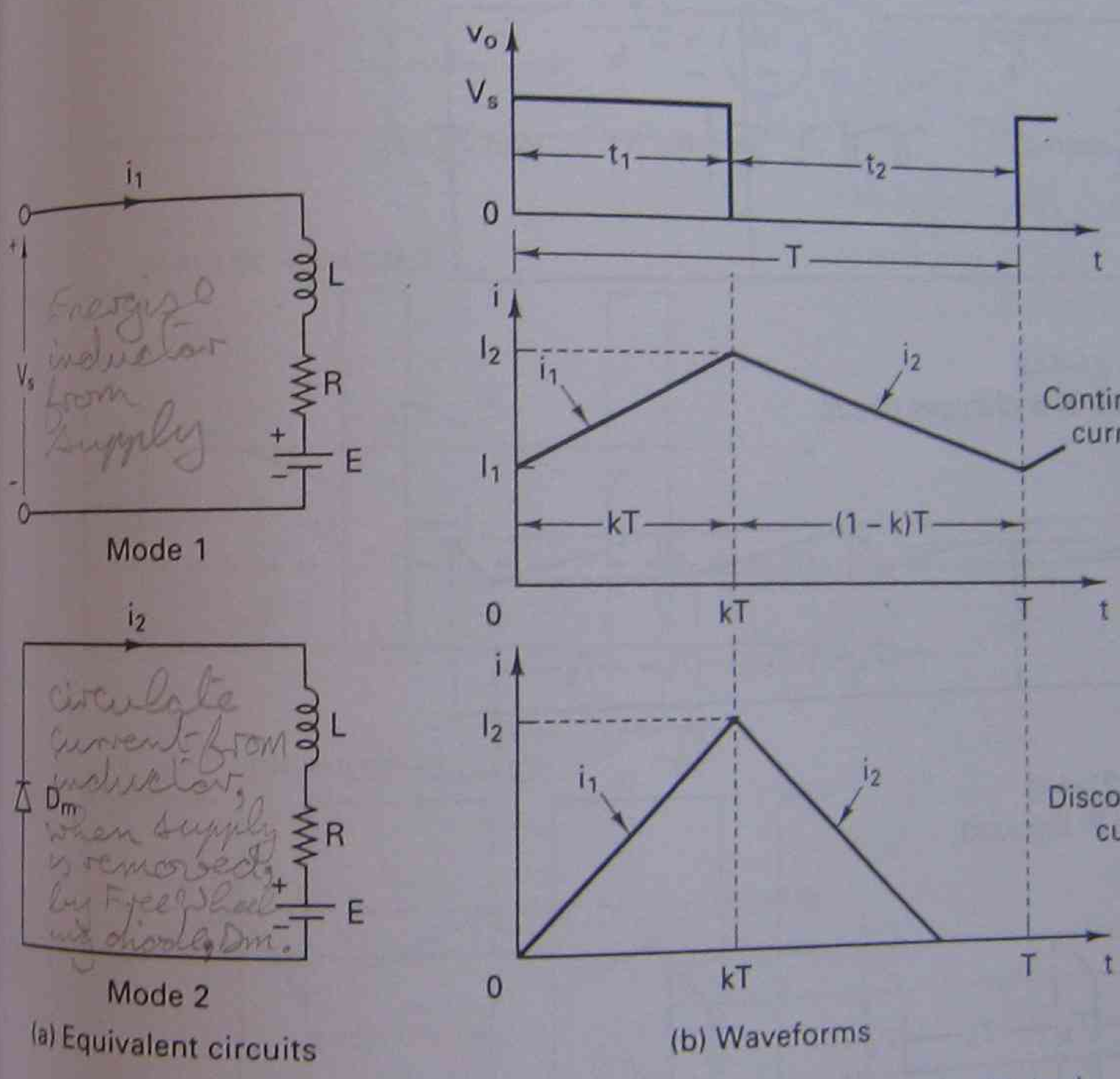
Energy stored in L

$W = \frac{1}{2}LI_a^2$

$(E_s - E_0)T_1 = E_0T_2$ -

When the chopper is turned on, the energy is transferred from the source V_s to inductor L. If the chopper is then turned off, a magnitude of the energy stored in the inductor is forced to battery E.

Note. Without the chopping action, v_s must be greater than E for transferring power from V_s to E.



(a) Equivalent circuits

(b) Waveforms

Figure 9-3 Equivalent circuits and waveforms for *RL* loads.

The power semiconductor devices require a minimum time to turn on and turn off. Therefore, the duty cycle k can only be controlled between a minimum value k_{min} and a maximum value k_{max}, thereby limiting the minimum and maximum value of output voltage. The switching frequency of the chopper is also limited. It can be noticed from Eq. (9-20) that the load ripple current depends inversely on the chopping frequency f. The frequency should be as high as possible to reduce the load ripple current and to minimize the size of any additional series inductor in the load circuit.

Eq(9-20) $\Delta I_{RIPPLE} = \frac{V_s}{4fL}$

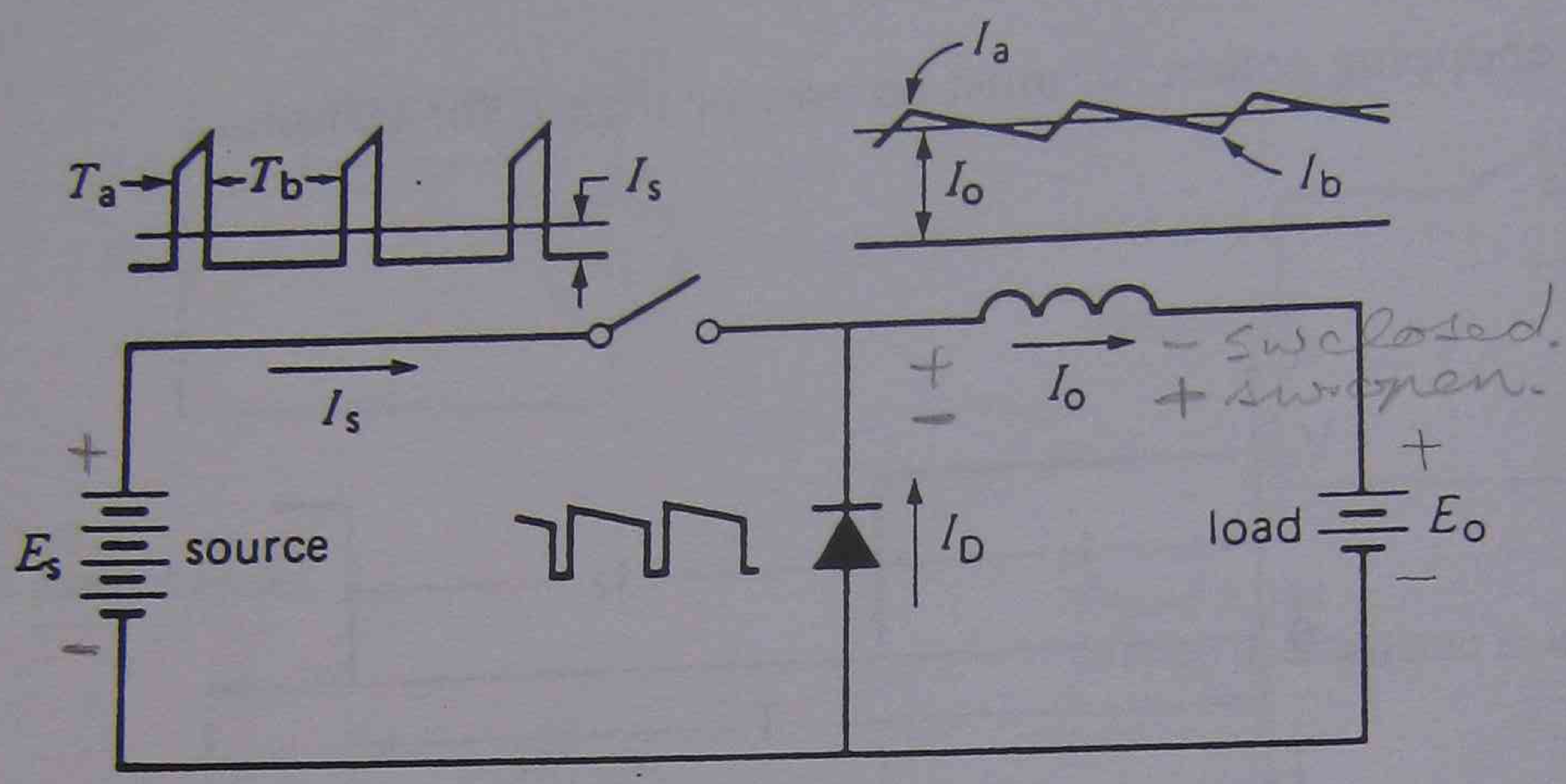


Figure 21-44a
Currents in a chopper circuit.

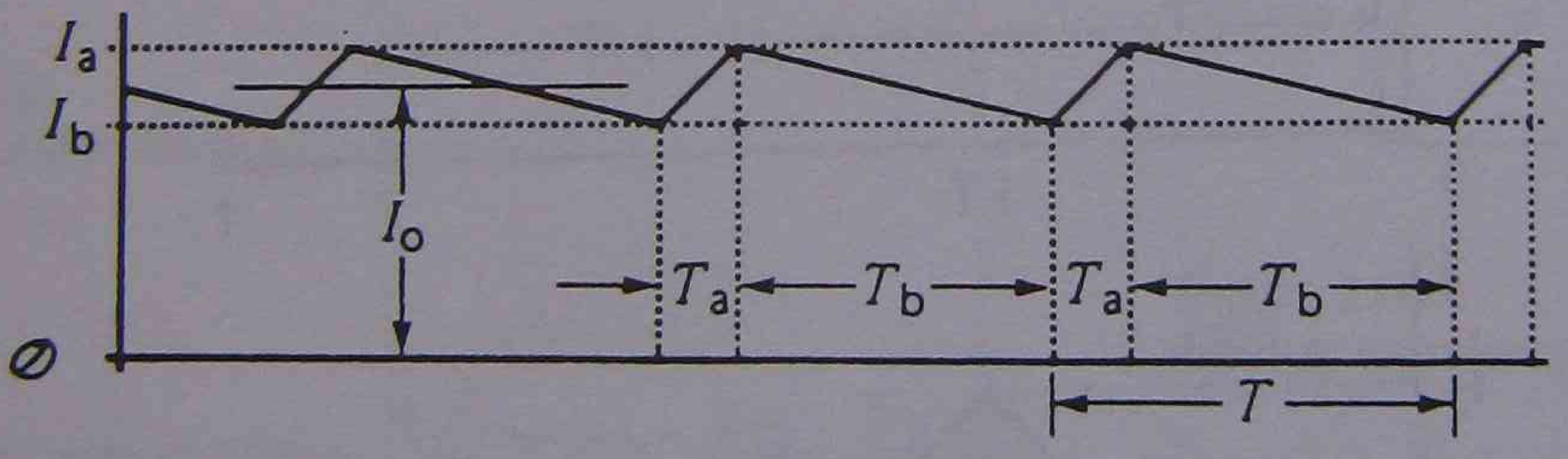


Figure 21-44b
Current in the load.

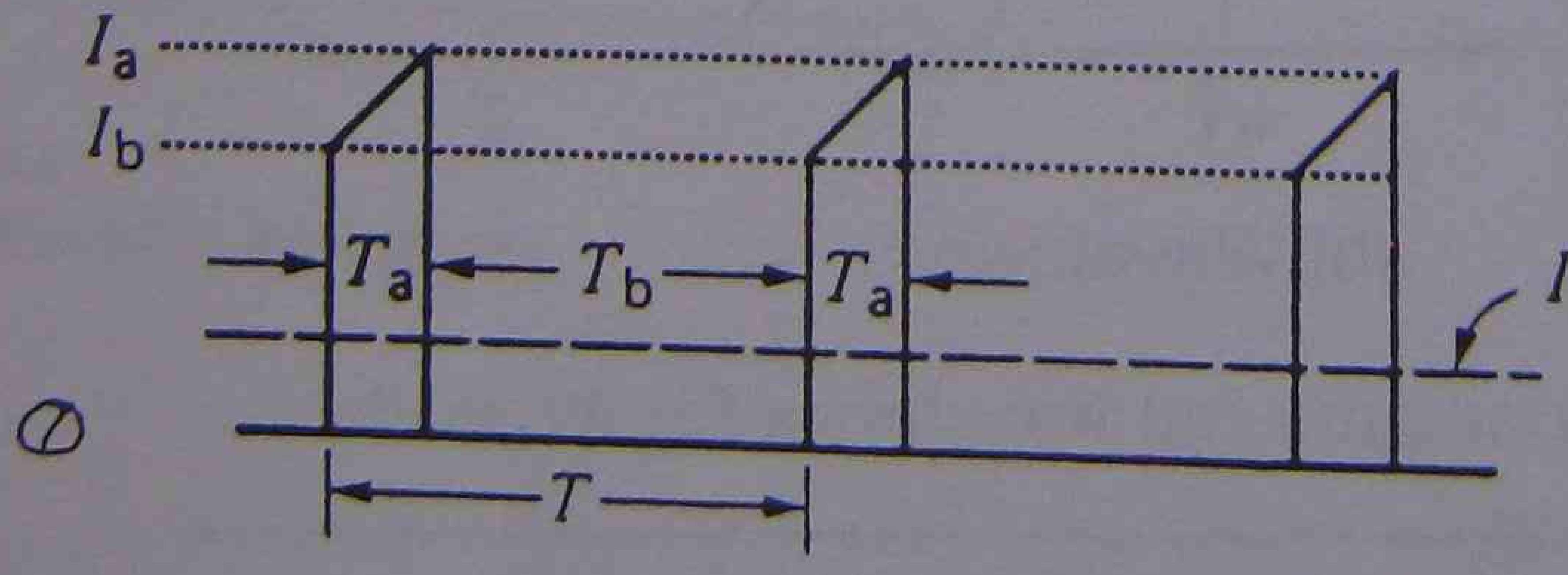


Figure 21-44c
Current drawn from the source.

A.C. Motor Control.

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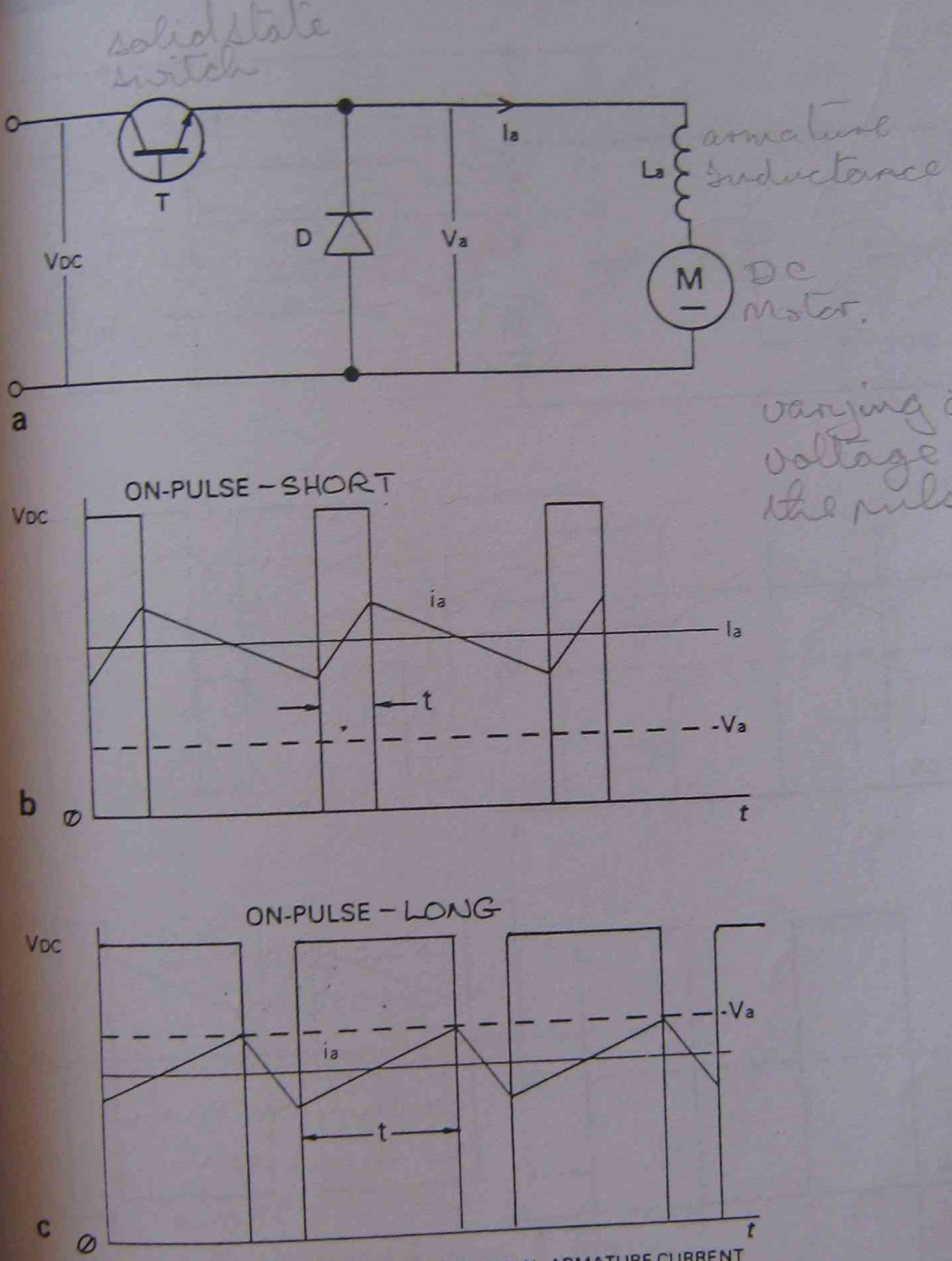

varying of applied voltage by varying the pulse width

VDC = DC SUPPLY VOLTAGE
Va = DC ARMATURE VOLTAGE
La = ARMATURE INDUCTANCE
Ia = MEAN ARMATURE CURRENT
ia = INSTANTANUOUS ARMATURE CURRENT

2 *Basic dc dc converter, or chopper:* **a** *circuit;* **b** *low output voltage;* **c** *high output voltage.*

A.C. Motor Control.

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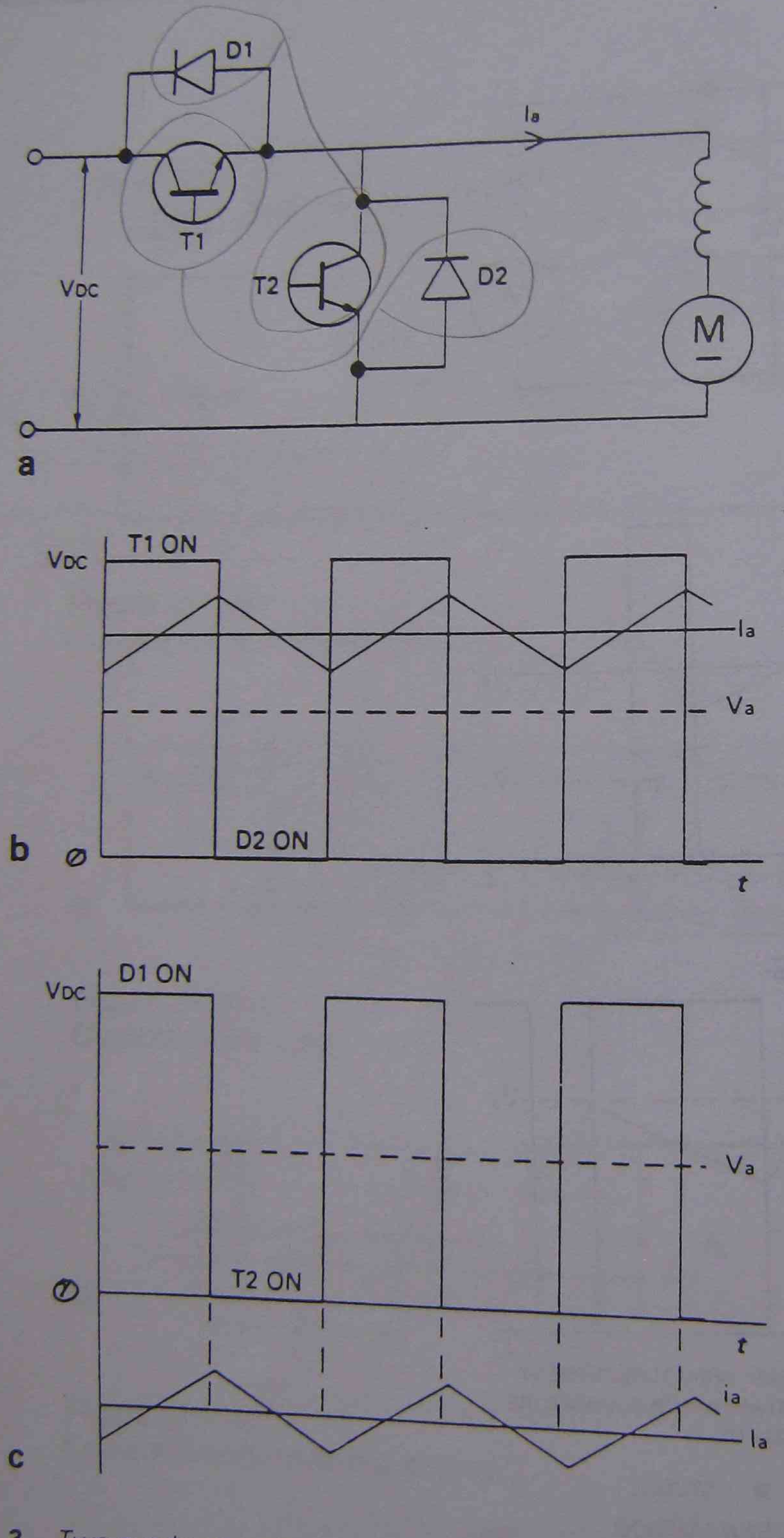

3 *Two-quadrant dc dc converter;* **a** *circuit;* **b** *forward motoring;* **c** *forward braking.*

2quadrant operation
Forward
- Motoring &,
- Braking.

- Quadrant 1
- Motoring
- Energy transferd from supply to motor

T1 & D2 control

- Quadrant 2
- Braking
- Energy transferd back from motor to the supply, i.e. the motor is a generator with the supply as the load.

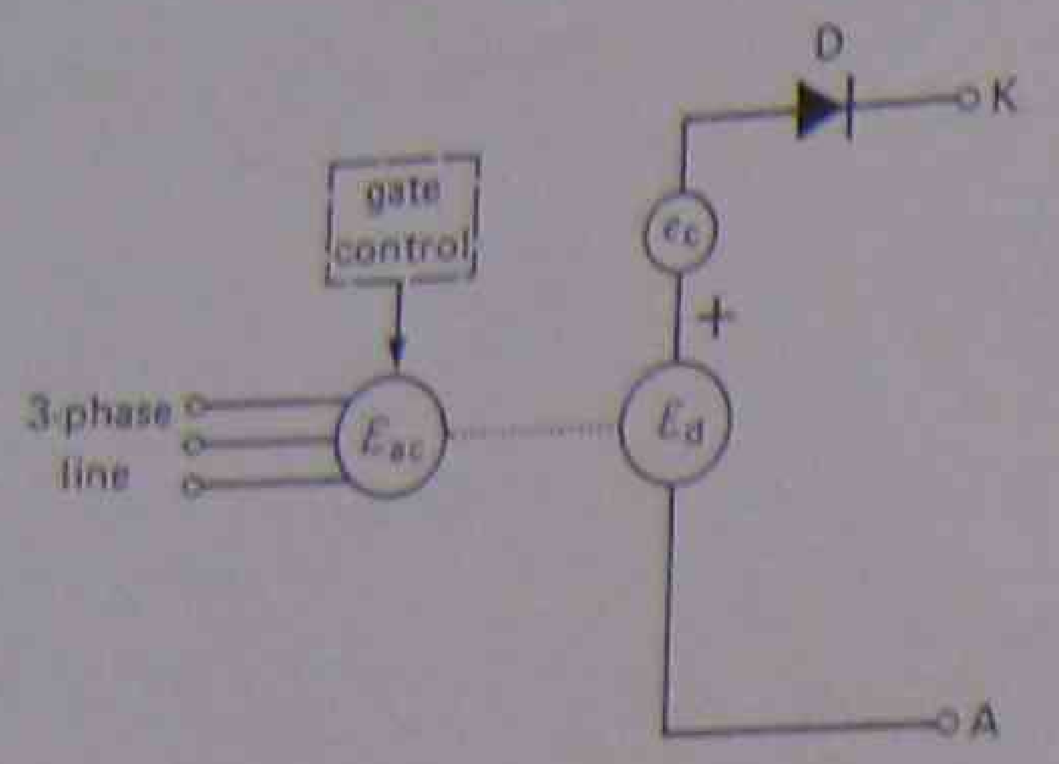

Figure 21-55
Equivalent circuit of a converter.

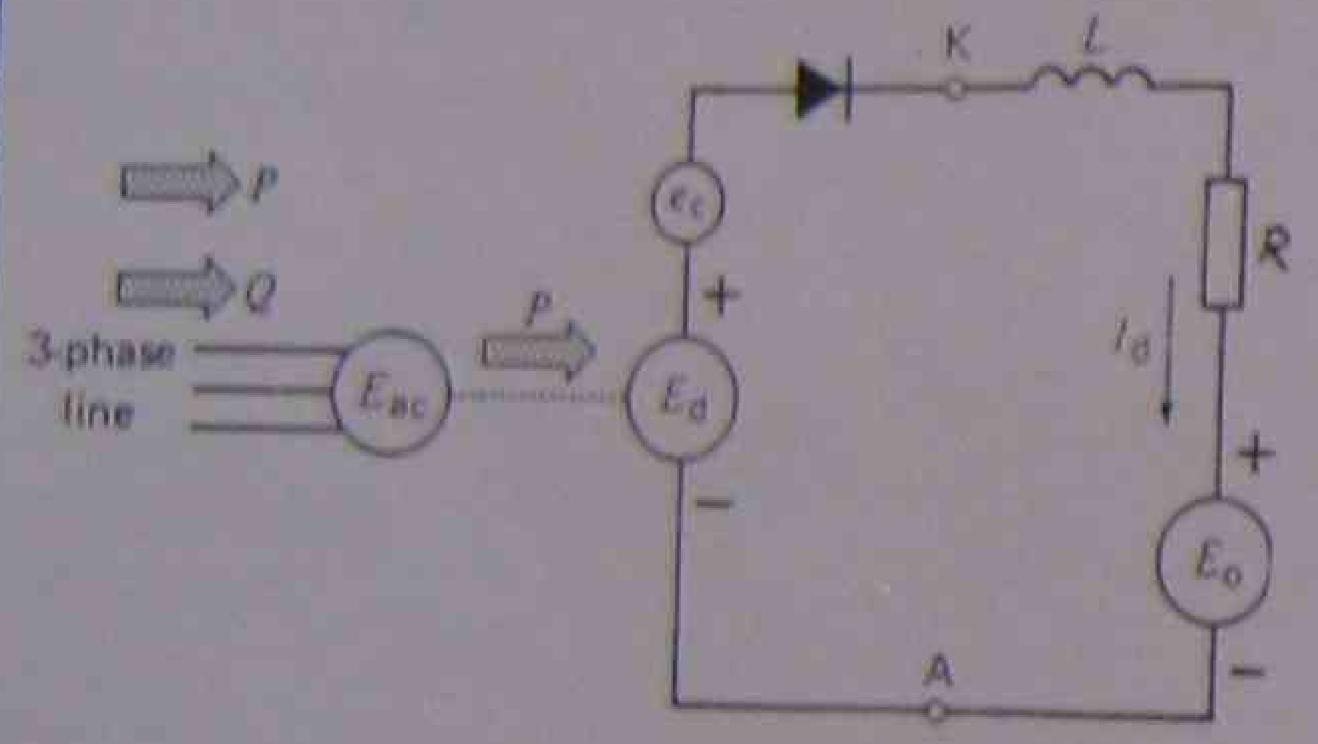

Figure 21-56
Equivalent circuit of a 3-phase converter in the rectifier mode.

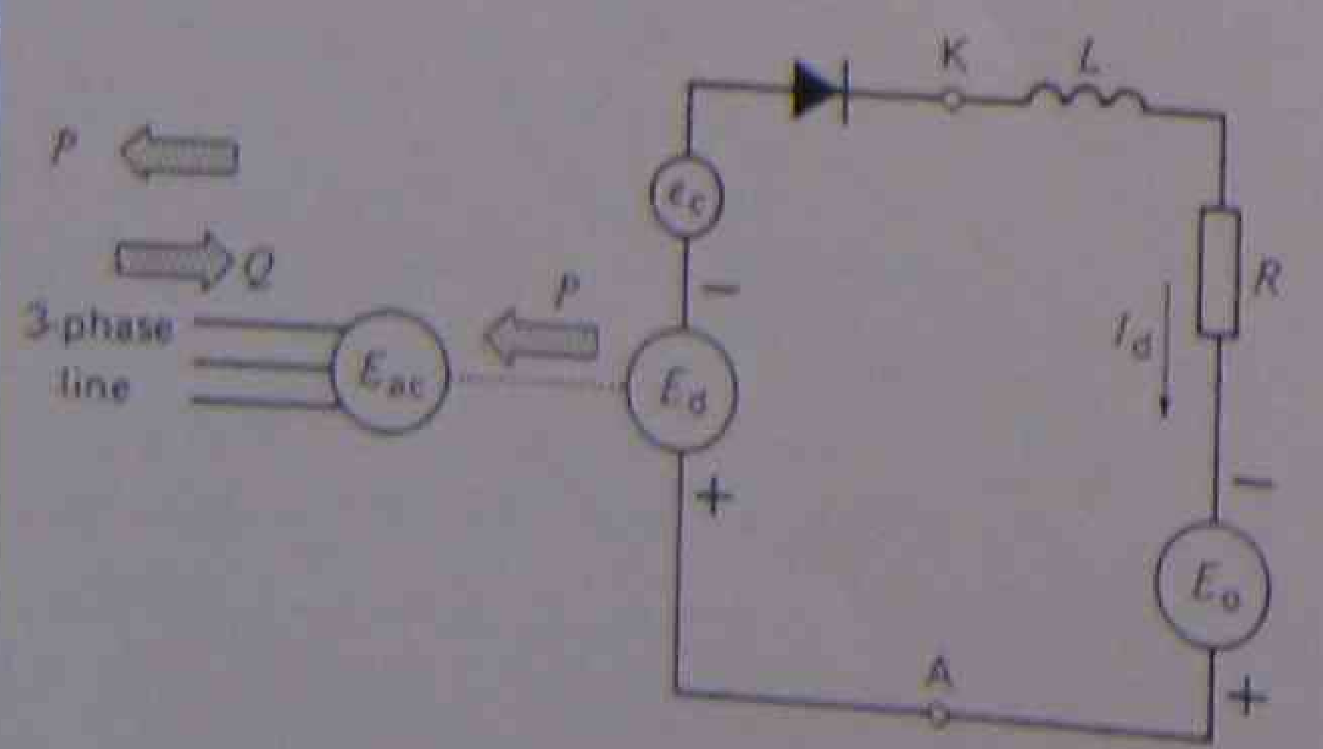

Figure 21-57
Equivalent circuit of a 3-phase converter in the inverter mode.

21.31 Equivalent circuit of a converter

We may think of a converter as being a st ac/dc motor-generator set whose dc output v tage E_d changes both in magnitude and pola depending upon the gate pulse delay. Ho ever, the dc "generator" has some spe properties:

1. it can carry current in only one direction
2. it generates an increasingly large ac ripp voltage as the dc voltage decreases.

The analogy may be represented by the c cuit of Fig. 21-55, in which:

- E_{ac} represents the 3-phase line voltage
- E_d is the dc voltage generated by the co verter
- e_c is the ac voltage generated by the co verter (mainly 6th and 12th harmonics)
- D is a diode to remind us that current c flow in only one direction
- The dotted line between E_{ac} and E_d indicate that active power can flow between the a and dc systems.
- Unlike a motor/generator set, the dc and a systems are not electrically isolated fro each other.

When the converter is operating as a re fier, the equivalent circuit is shown Fig. 21-56. When operating as in inverter circuit is given by Fig. 21-57. The ac vo generated by the converter appears ac inductor L. Its inductance is sufficiently larg ensure an almost ripple-free dc current.

$$Q = P \tan \alpha \qquad (21\text{-}18$$

where

Q = reactive power absorbed by the converter [var]
P = dc power of the converter (positive for a rectifier, negative for an inverter) [W]
α = triggering angle [°]

21.32 Currents in a 3-phase, 6-pulse converter

Figure 21-58 shows the voltage and current waveshapes when the converter functions as a rectifier at a firing angle of 45°. The current i_1, i_2, i_3, i_4, i_5, i_6 in each thyristor flows for 120° and its peak value is equal to the dc current I_d This holds true for any firing angle between zero and 180°. Consequently, the currents in a thyristor converter are identical to those in a plain 3-phase diode rectifier (Fig. 21-20). The only difference is that they flow later in the cycle.

The waveshapes of the corresponding ac line currents are easily found because they are equal to the difference between the respective thyristor currents. Thus, referring to Fig. 21-52, line current $I_a = i_1 - i_4$. These line currents also have a peak value I_d, but they flow in positive and negative pulses of 120°.

21.33 Power factor

When the currents are in phase with the voltages, the so-called *displacement power factor* is 100%. As a result, the rectifier draws no reactive power from the line. The same remarks apply to a 3-phase, 6-pulse rectifier (Fig. 21-20)

Referring now to Fig. 21-58, where triggering has been delayed by 45°, we note that the thyristor currents have all been shifted ("displaced") by 45°, to the right. Consequently, the line currents lag the respective voltages by 45°; the displacement power factor is no longer unity but only 0.707 (cos 45° = 0.707). This means that a converter absorbs reactive power from the ac system to which it is connected. This is true whether the converter operates as a rectifier or inverter. The reactive power is given by:

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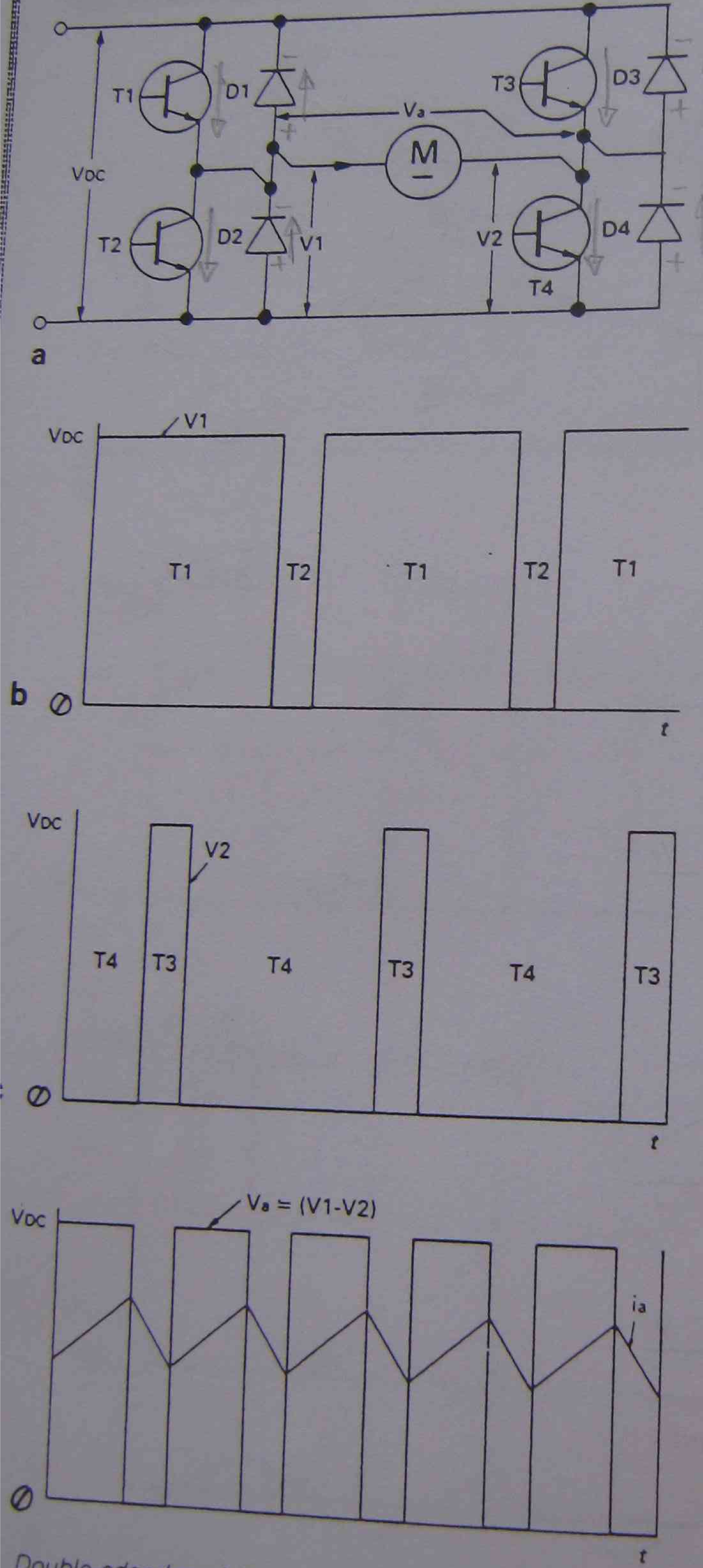
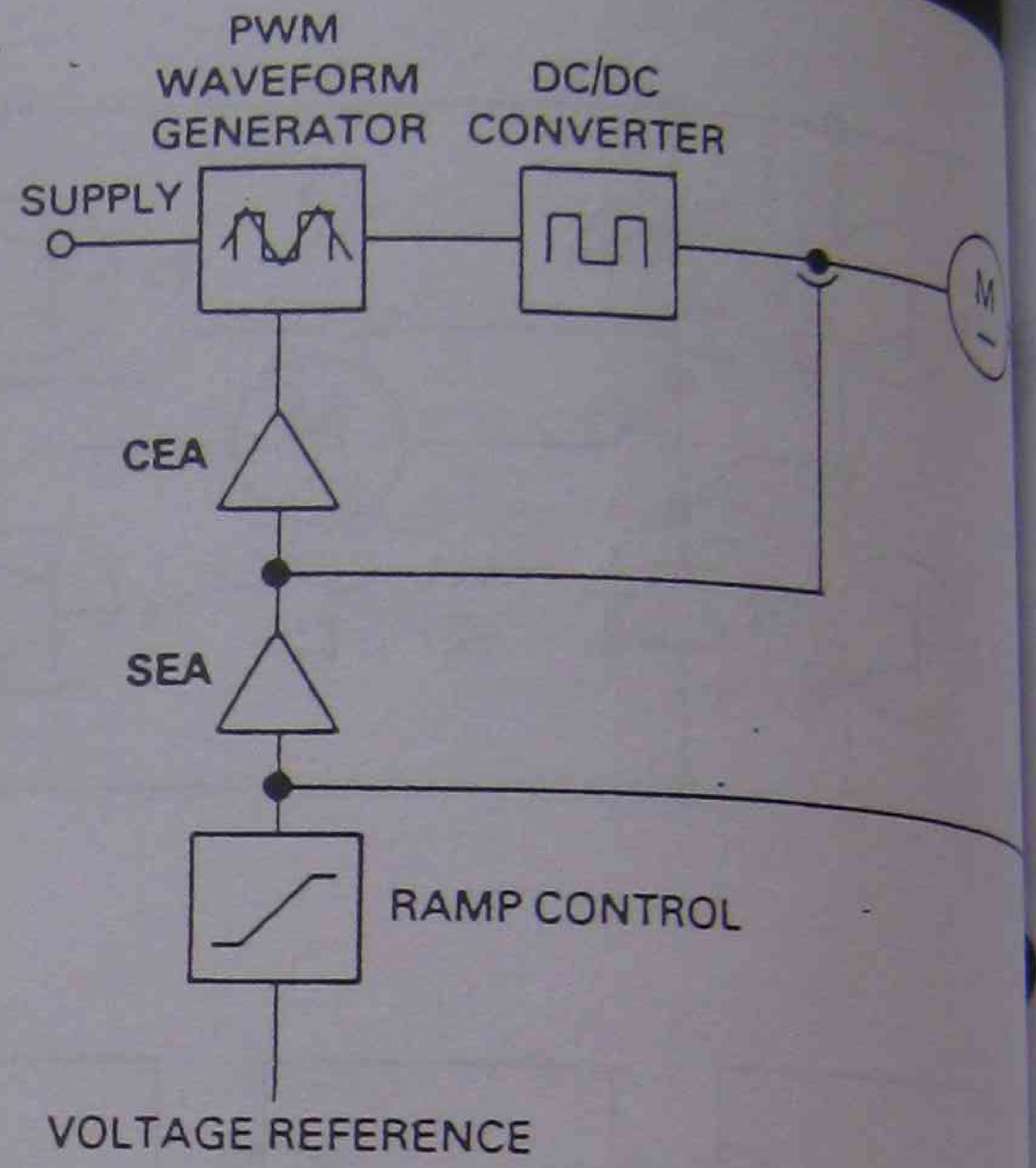
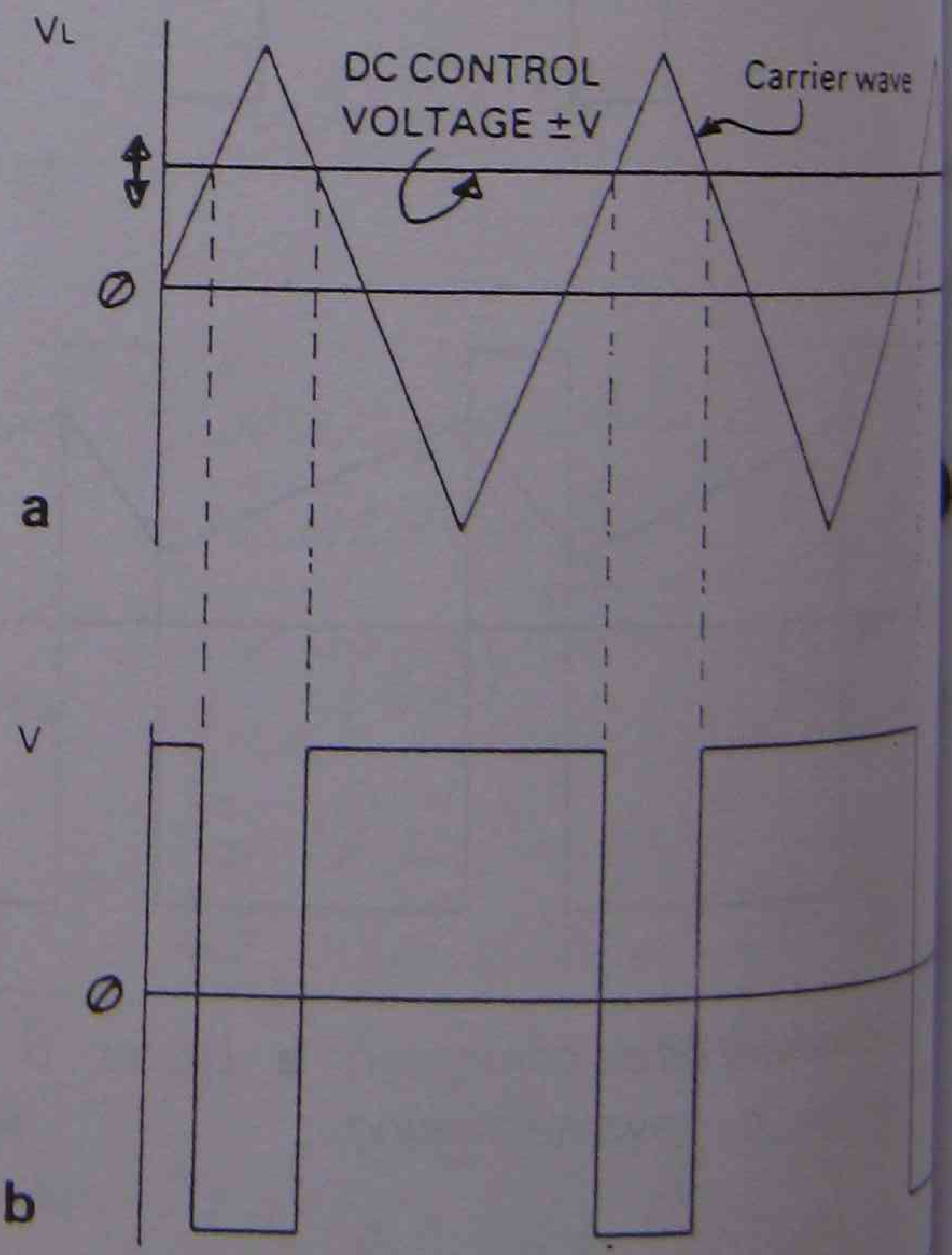


Double-edged modulation; **a** circuit; **b** conduction periods of T1 and T2; **c** conduction periods of T3 and T4; **d** resultant output.



6 *Typical transistor dc drive, control loops.*



7 *PWM waveform generation;* **a** *carrier wav control voltage;* **b** *converter transistor cont waveform.*

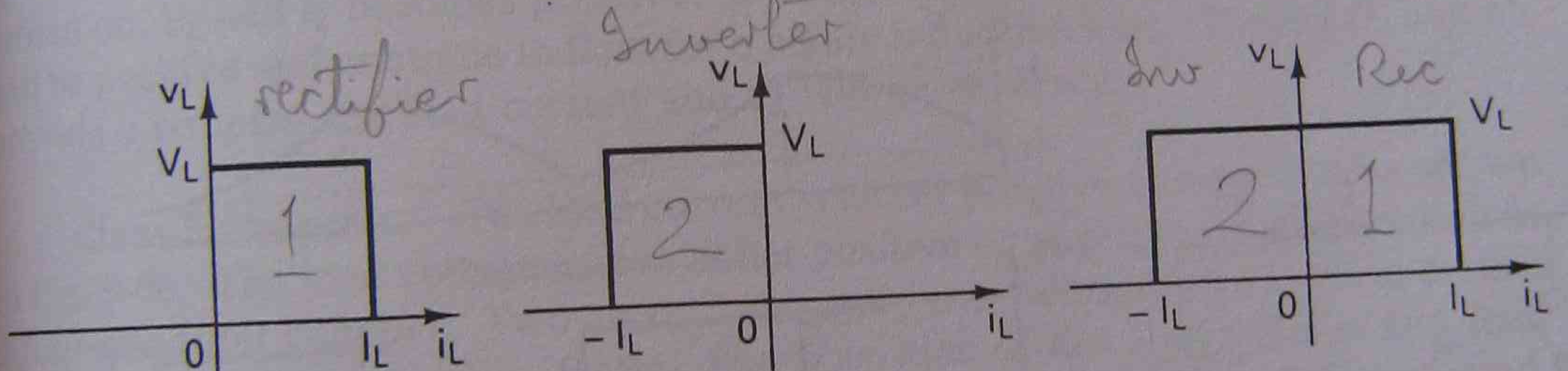
8

A.C. Motor Control.

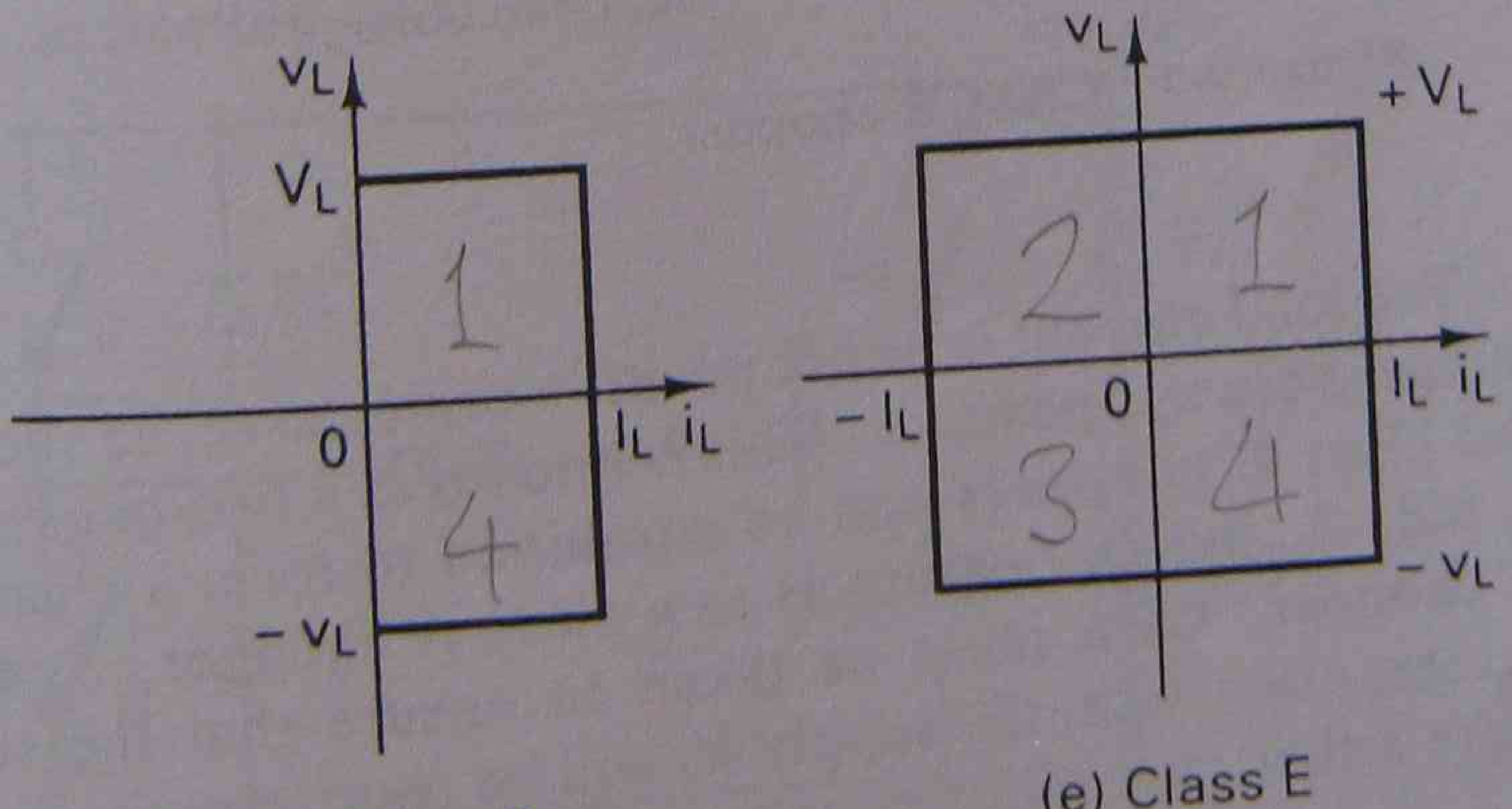
Topic -

The step-down chopper in Fig. 9-1a only allows power to flow from the supply to the load, and is referred to as class A chopper. Depending on the directions of current and voltage flows, choppers can be classified into five types:

Class A chopper
Class B chopper
Class C chopper
Class D chopper
Class E chopper



(a) Class A
(b) Class B
(c) Class C



(d) Class D
(e) Class E

Figure 9-6 Chopper classification.

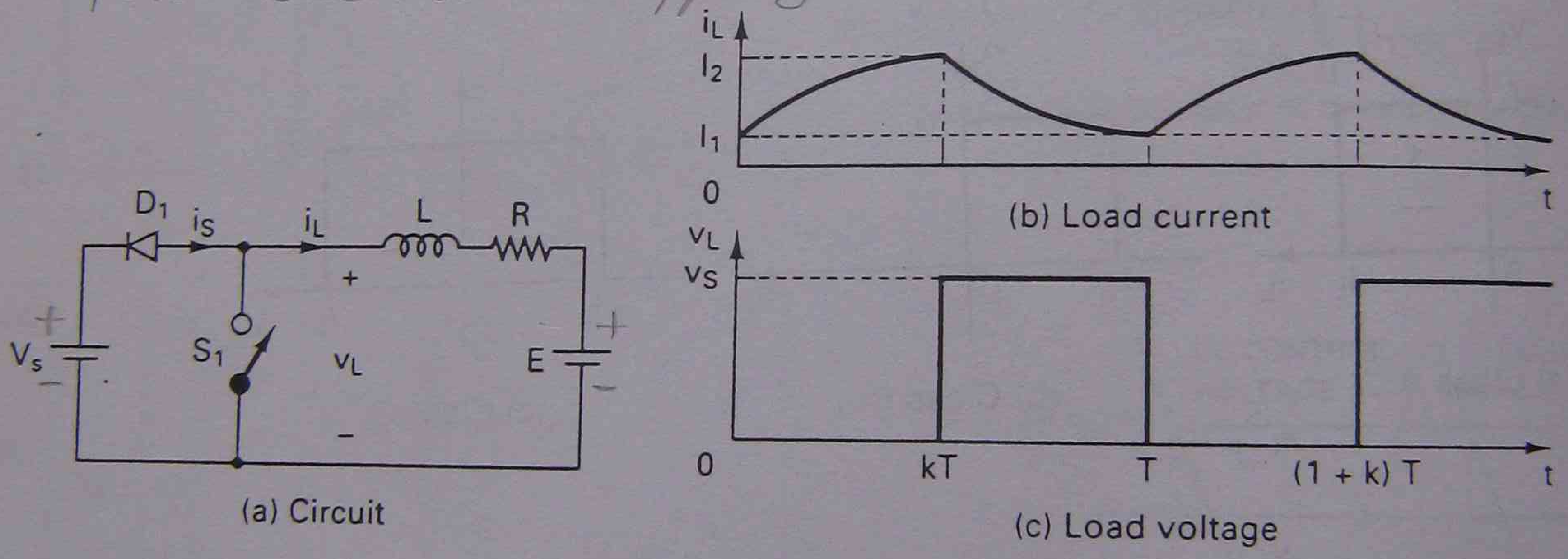
A.C. Motor Control.

Topic -

Class A chopper. The load current flows into the load. Both the load voltage and the load current are positive, as shown in Fig. 9-6a. This is a single-quadrant chopper and is said to be operated as a rectifier.

Class B chopper. The load current flows out of the load. The load voltage is positive, but the load current is negative, as shown in Fig. 9-6b. This is also a single-quadrant chopper, but operates in the second quadrant and is said to be operated as an inverter. A class B chopper is shown in Fig. 9-7a, where the battery E is a part of the load and may be the back emf of a dc motor.

When switch S_1 is turned on, the voltage E drives current through inductor L and load voltage v_L becomes zero. The instantaneous load voltage v_L and load voltage E combine to forward bias D1 and supply power back to the supply.



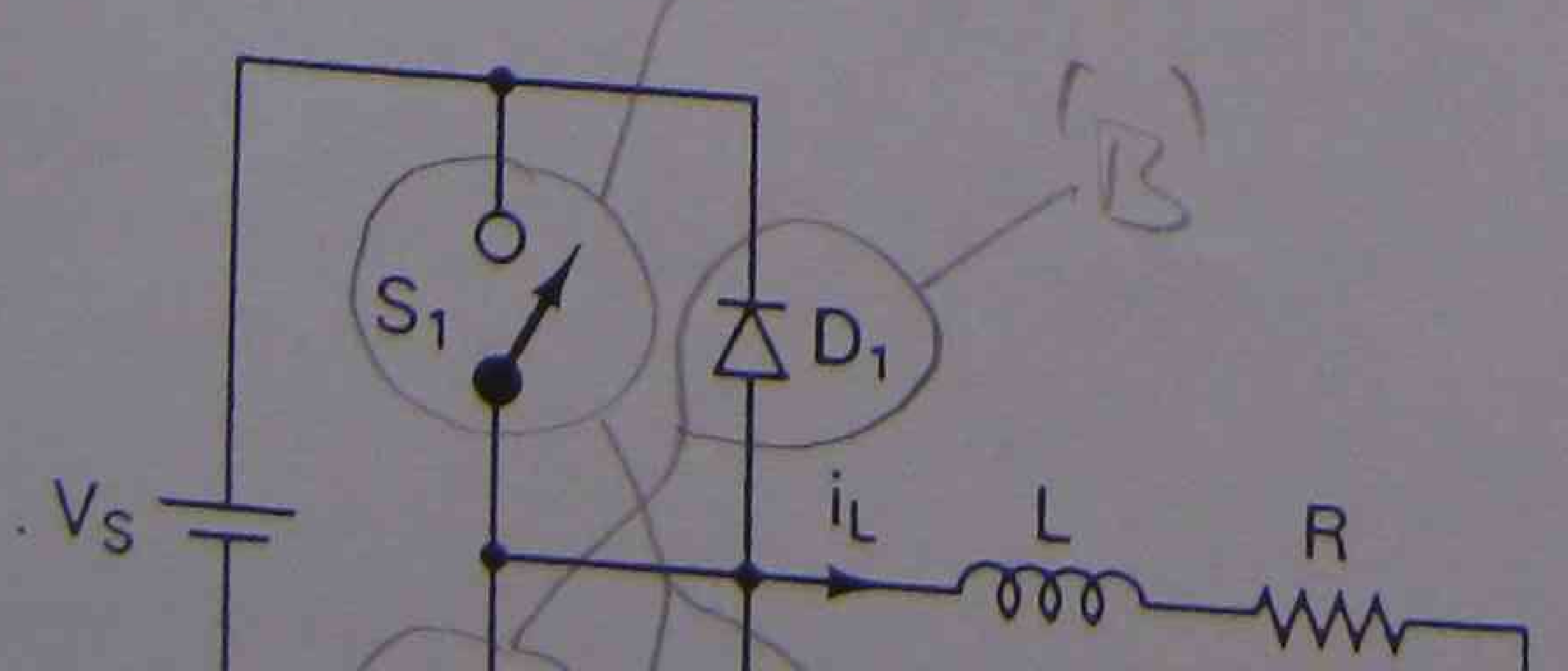
(a) Circuit

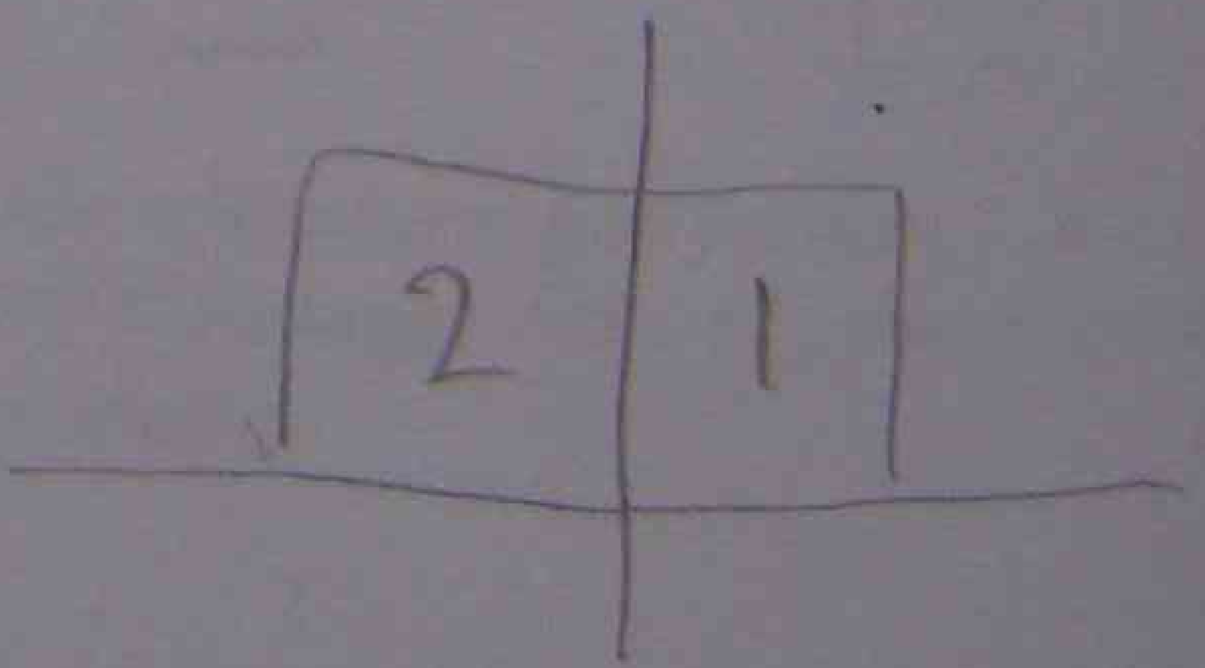
(b) Load current

(c) Load voltage

Figure 9-7 Class B chopper.

Class C chopper. The load current is either positive or negative, as shown in Fig. 9-6c. The load voltage is always positive. This is known as a *two-quadrant chopper*. The class A and class B choppers can be combined to form a class C chopper as shown in Fig. 9-8. S_1 and D_2 operate as as a class A chopper. S_2 and D_1 operate as a class B chopper. Care must be taken to ensure that the two switches are not fired together; otherwise, the supply V_s will be short-circuited. A class C chopper can operate either as a rectifier or as an inverter.

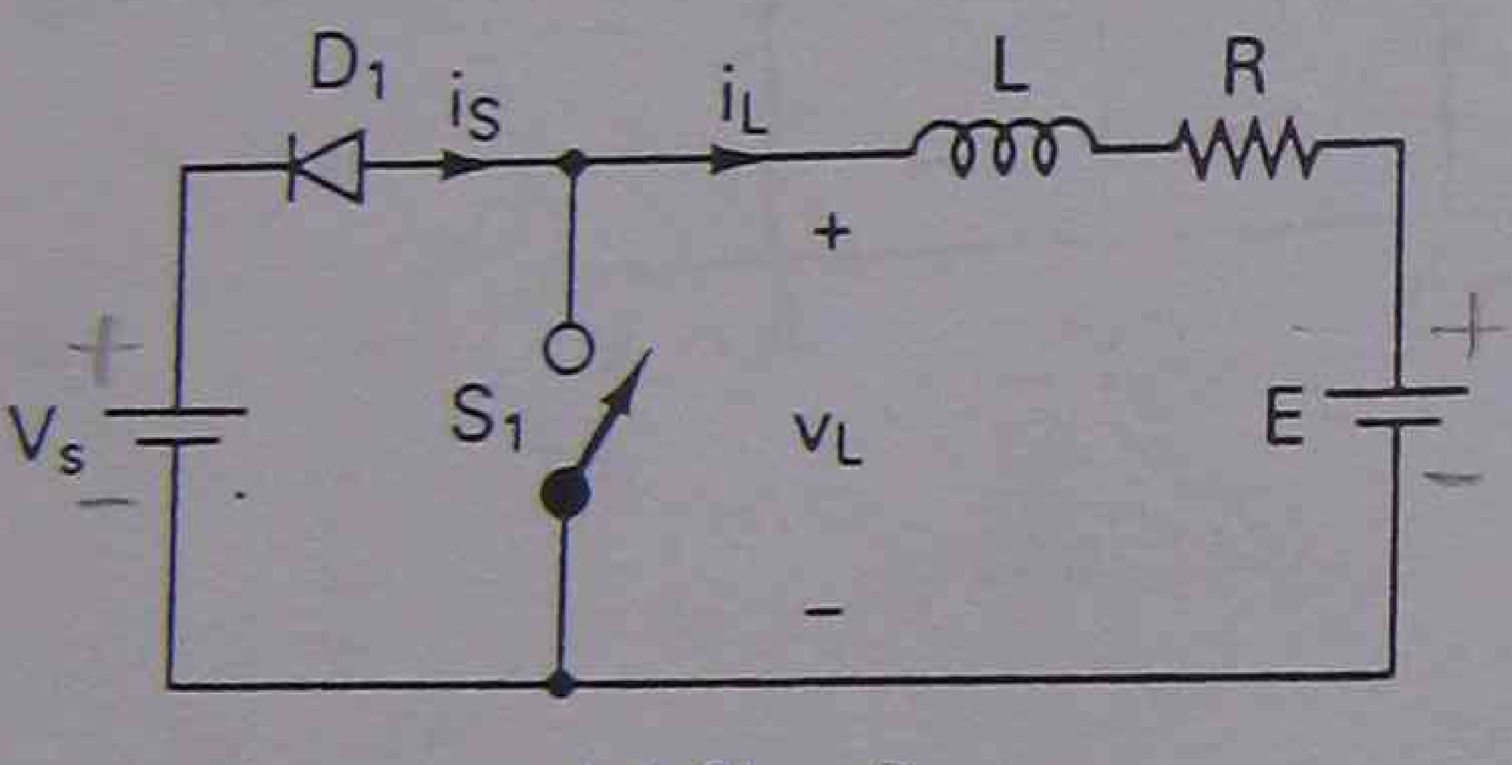




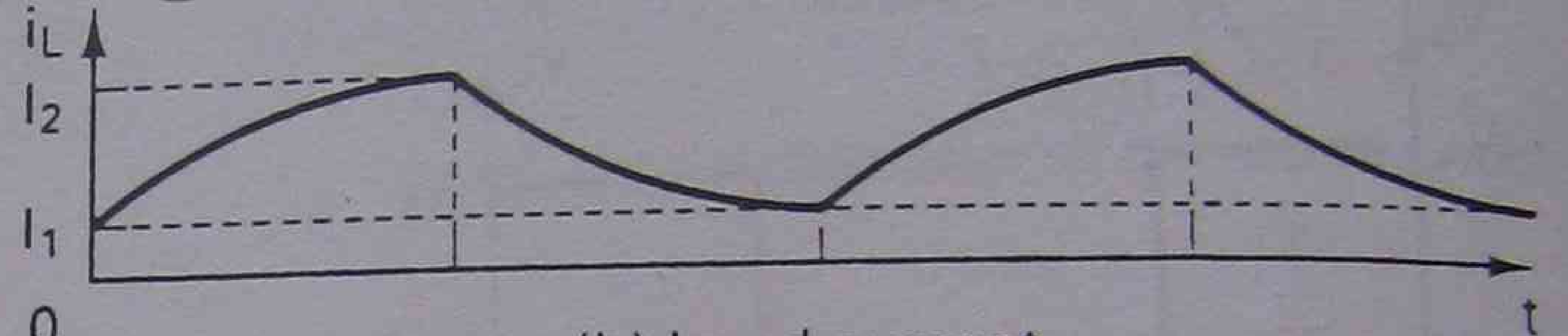
Class A chopper. The load current flows into the load. Both the load voltage and the load current are positive, as shown in Fig. 9-6a. This is a single-quadrant chopper and is said to be operated as a rectifier.

Class B chopper. The load current flows out of the load. The load voltage is positive, but the load current is negative, as shown in Fig. 9-6b. This is also a single-quadrant chopper, but operates in the second quadrant and is said to be operated as an inverter. A class B chopper is shown in Fig. 9-7a, where the battery E is a part of the load and may be the back emf of a dc motor.

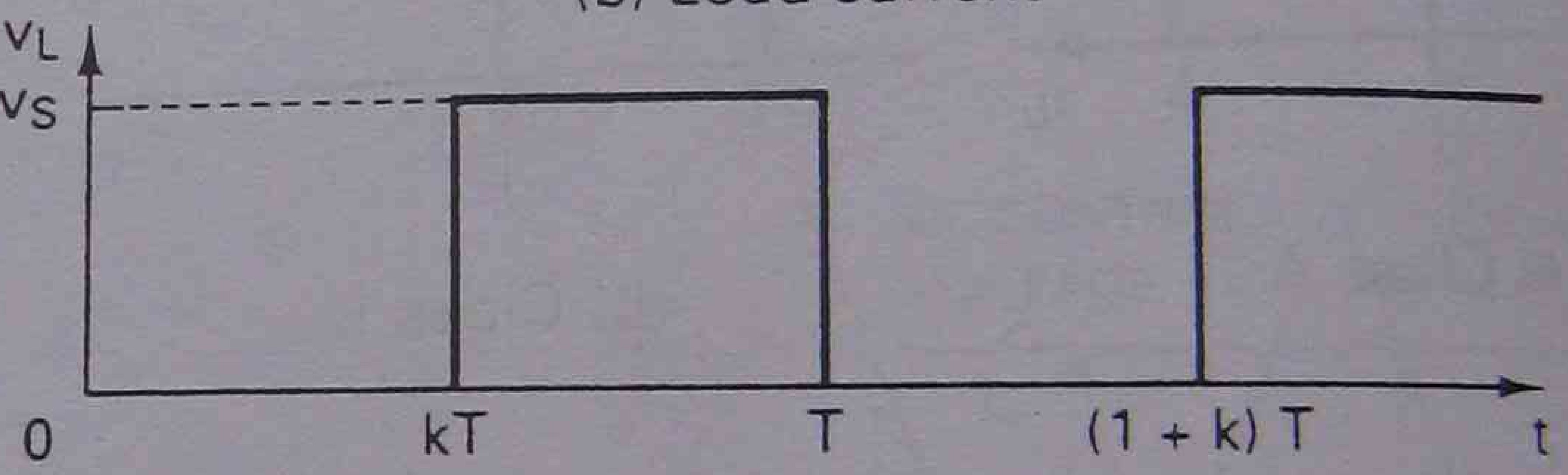
When switch S_1 is turned on, the voltage E drives current through inductor L and load voltage v_L becomes zero. The instantaneous load voltage v_L and load voltage E combine to forward bias D_1 and supply power back to the supply.



(a) Circuit



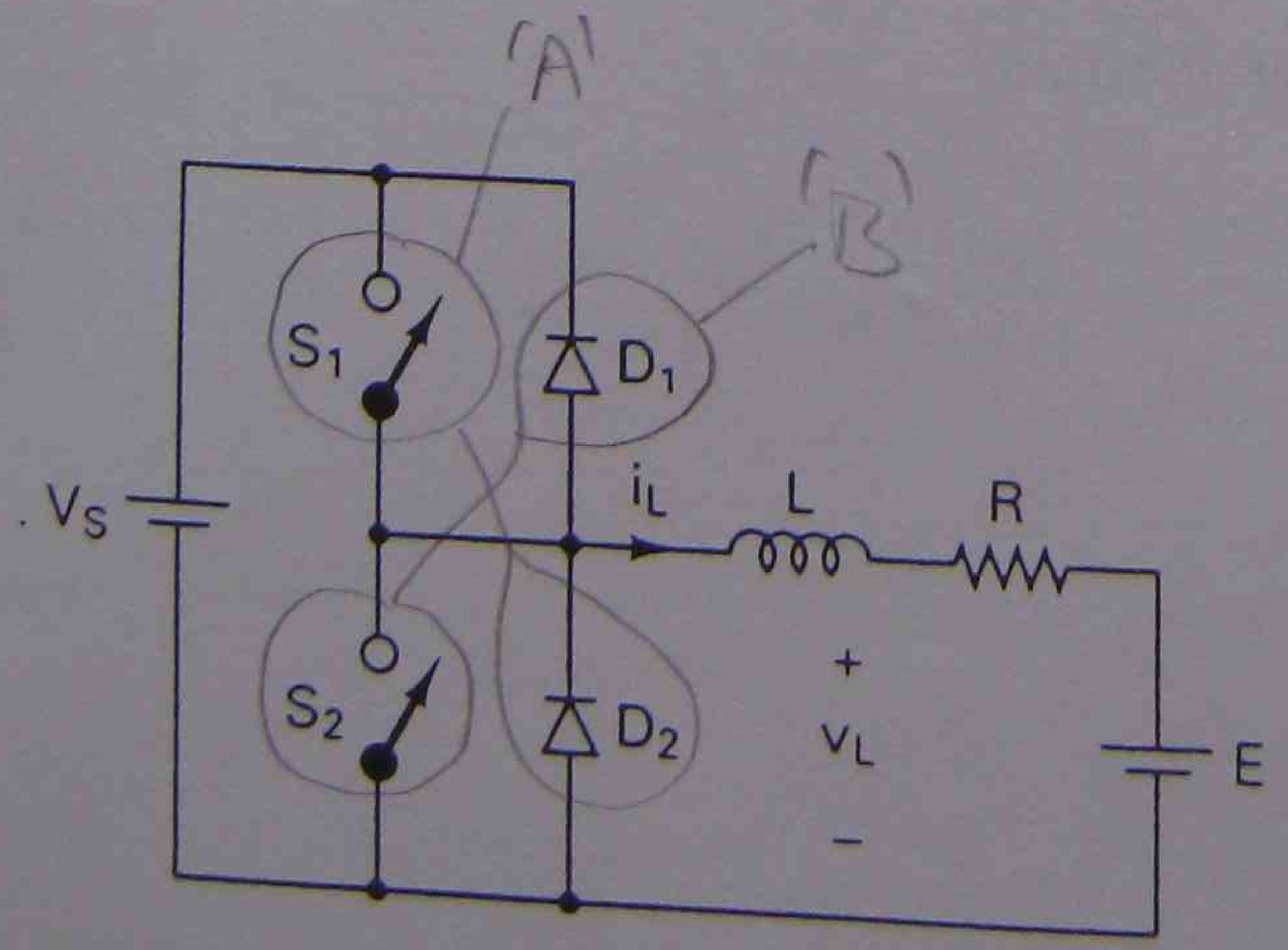
(b) Load current

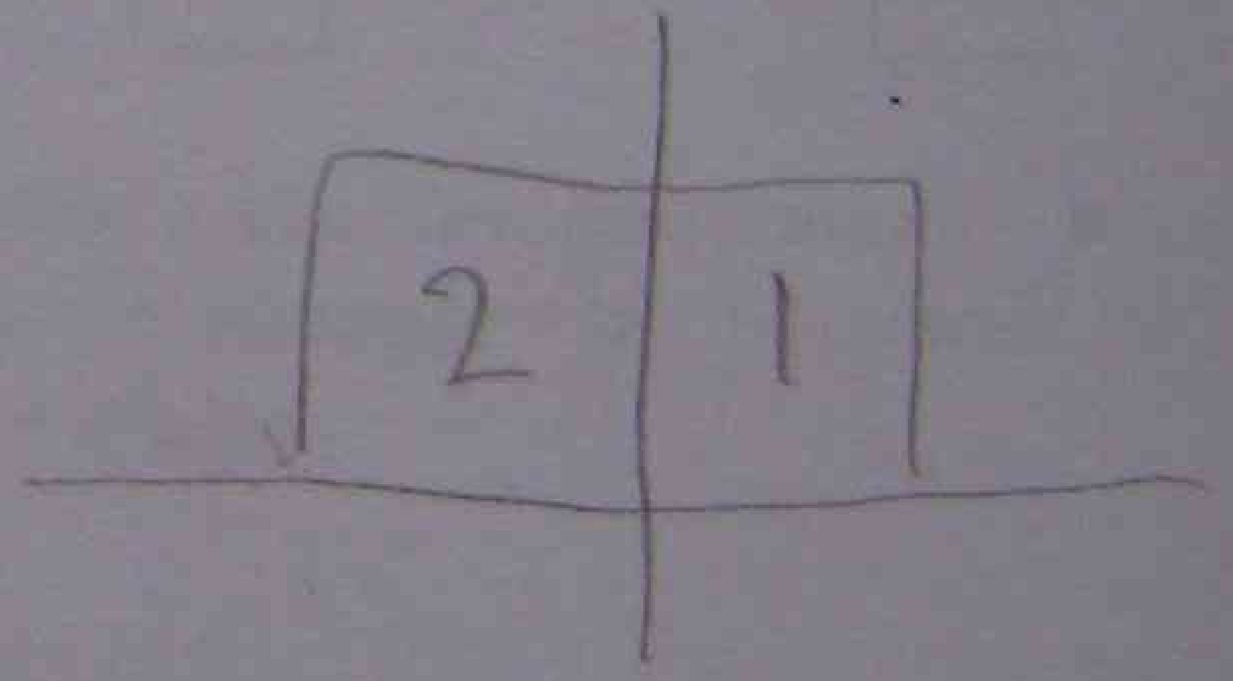


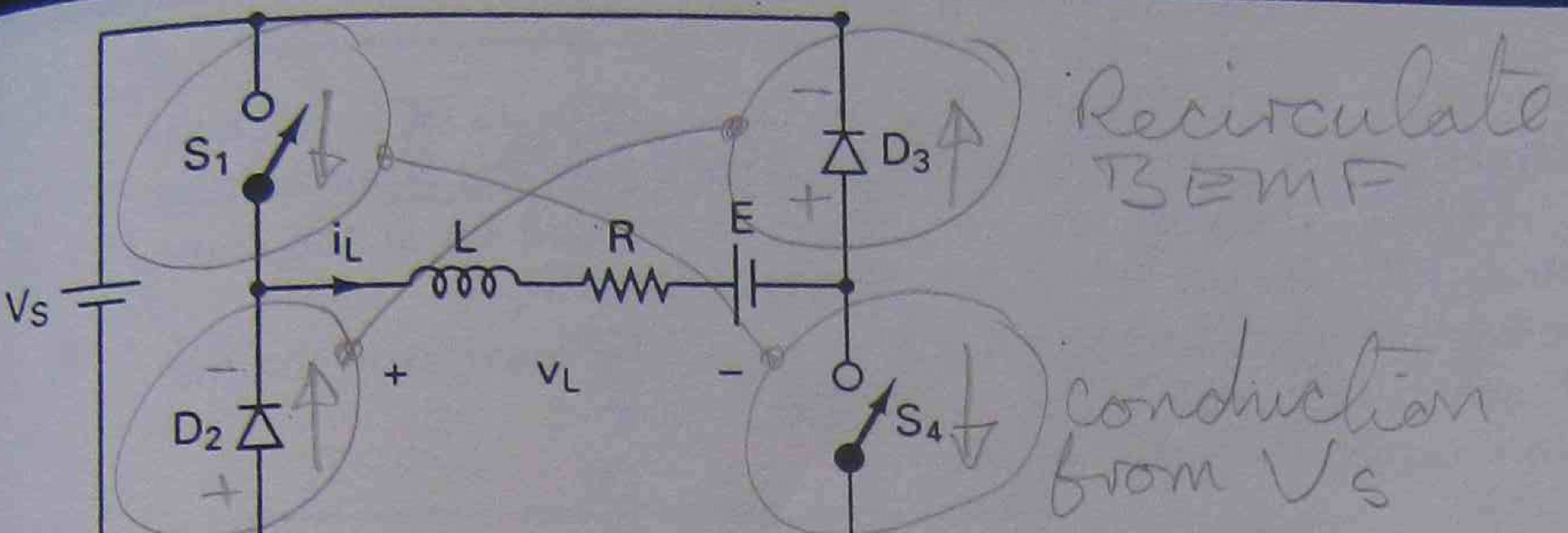
(c) Load voltage

Figure 9-7 Class B chopper.

Class C chopper. The load current is either positive or negative, as shown in Fig. 9-6c. The load voltage is always positive. This is known as a *two-quadrant chopper*. The class A and class B choppers can be combined to form a class C chopper as shown in Fig. 9-8. S_1 and D_2 operate as as a class A chopper. S_2 and D_1 operate as a class B chopper. Care must be taken to ensure that the two switches are not fired together; otherwise, the supply V_s will be short-circuited. A class C chopper can operate either as a rectifier or as an inverter.







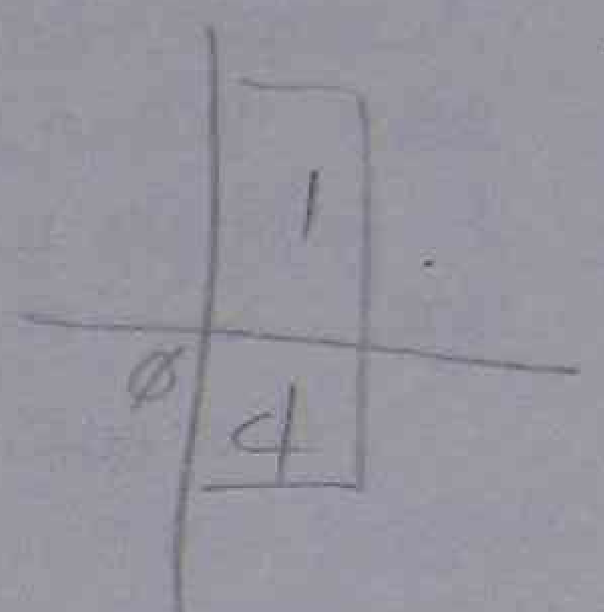
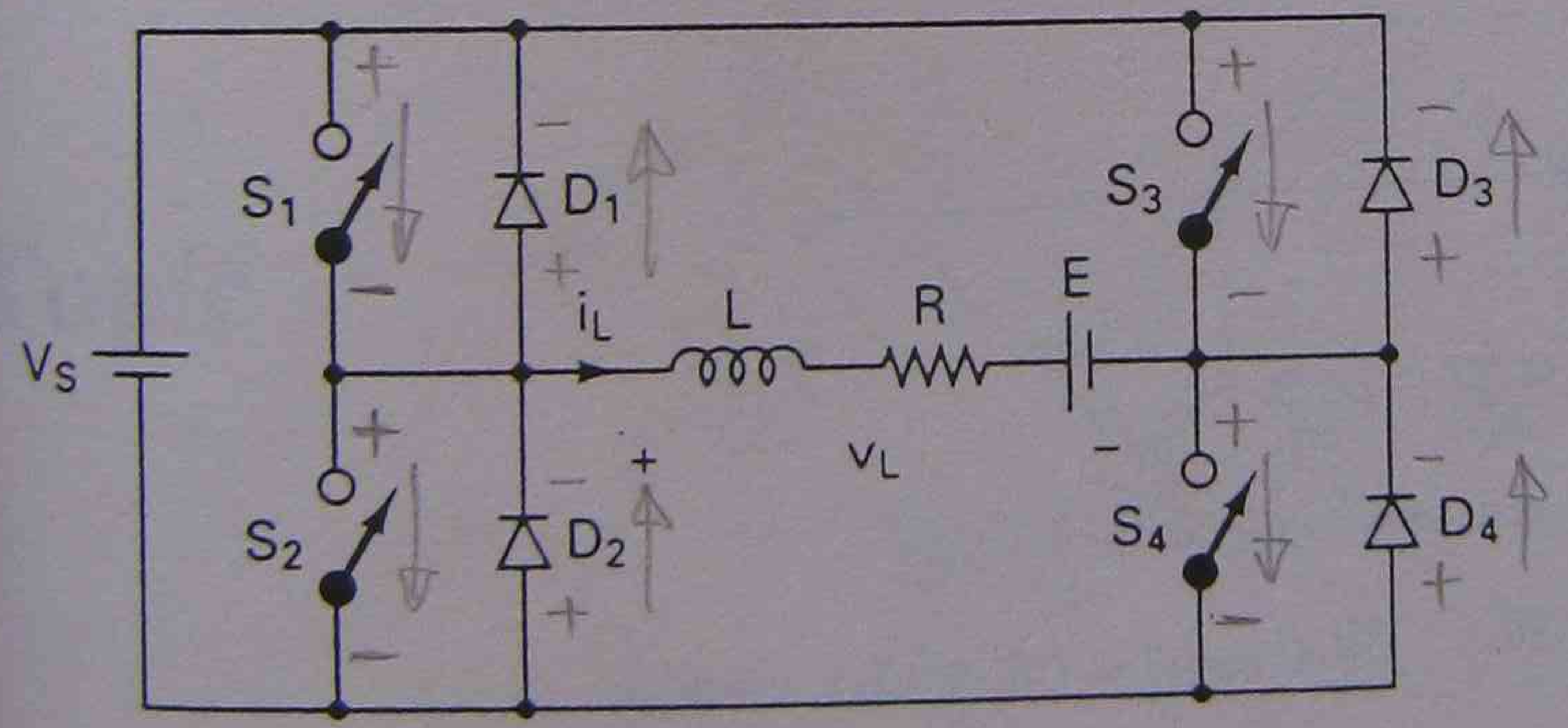


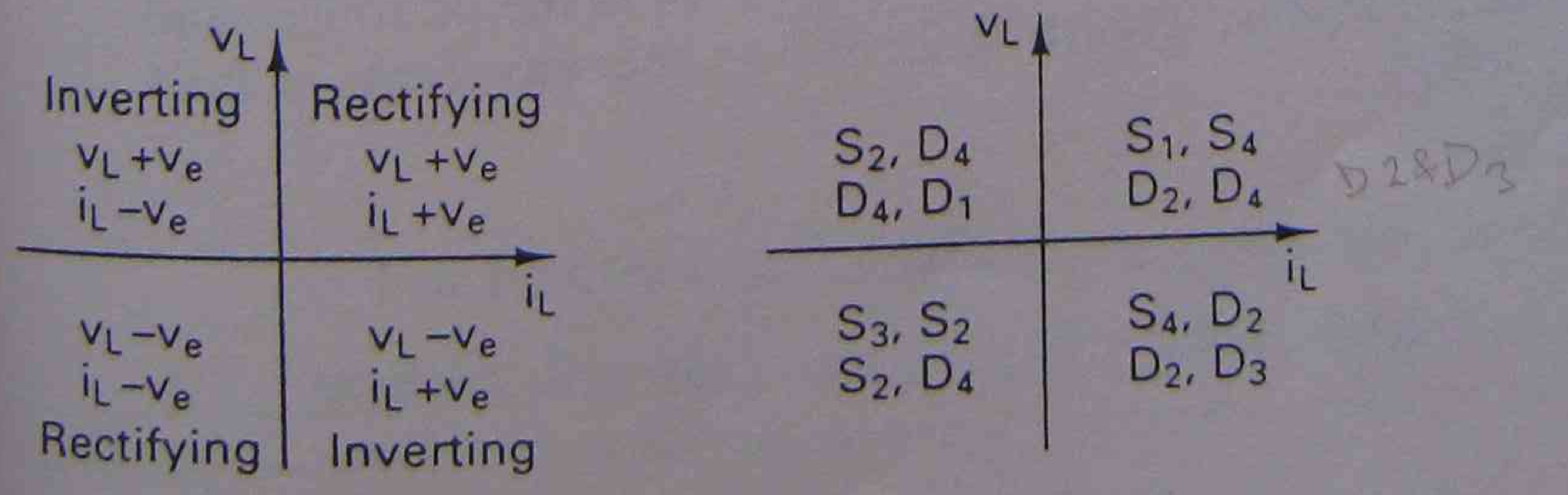
Figure 9-9 Class D chopper.

Class D chopper. The load current is always positive. The load voltage is either positive or negative, as shown in Fig. 9-6d. A class D chopper can also operate either as a rectifier or as an inverter, as shown in Fig. 9-9. If S_1 and S_4 are turned on, v_L and i_L becomes positive. If S_1 and S_4 are turned off, load current i_L will be positive and continue to flow for a highly inductive load. Diodes D_2 and D_3 provide a path for the load current and v_L will be reversed.

Class E chopper. The load current is either positive or negative, as shown in Fig. 9-6e. The load voltage is also either positive or negative. This is known as a *four-quadrant chopper*. Two class C choppers can be combined to form a class E chopper, as shown in Fig. 9-10a. The polarities of the load voltage and load current are shown in Fig. 9-10b. The devices that are operative in different quadrants are shown in Fig. 9-10c. For operation in the fourth quadrant, the direction of battery E must be reversed. This chopper is the basis for the single-phase full-bridge inverter in Section 10-4.



(a) Circuit



(b) Polarities (c) Conducting devices

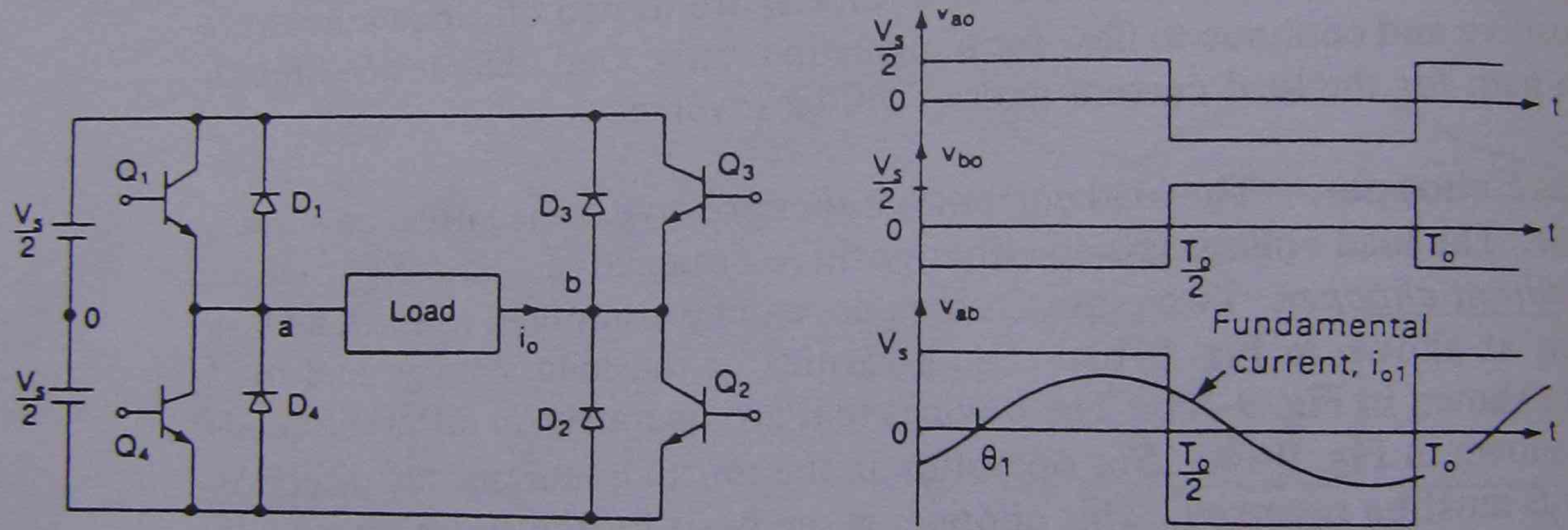
Figure 9-10 Class E chopper.

A single-phase bridge inverter is shown in Fig. 10-2a. It consists of four choppers. When transistors Q_1 and Q_2 are turned on simultaneously, the input voltage V_s appears across the load. If transistors Q_3 and Q_4 are turned on at the same time, WITH TRANSISTORS Q1 AND Q2 TURNED OFF, the voltage across the load is reversed and is $-V_s$. The waveform for the output voltage is shown in Fig. 10-2b.

The rms output voltage can be found from

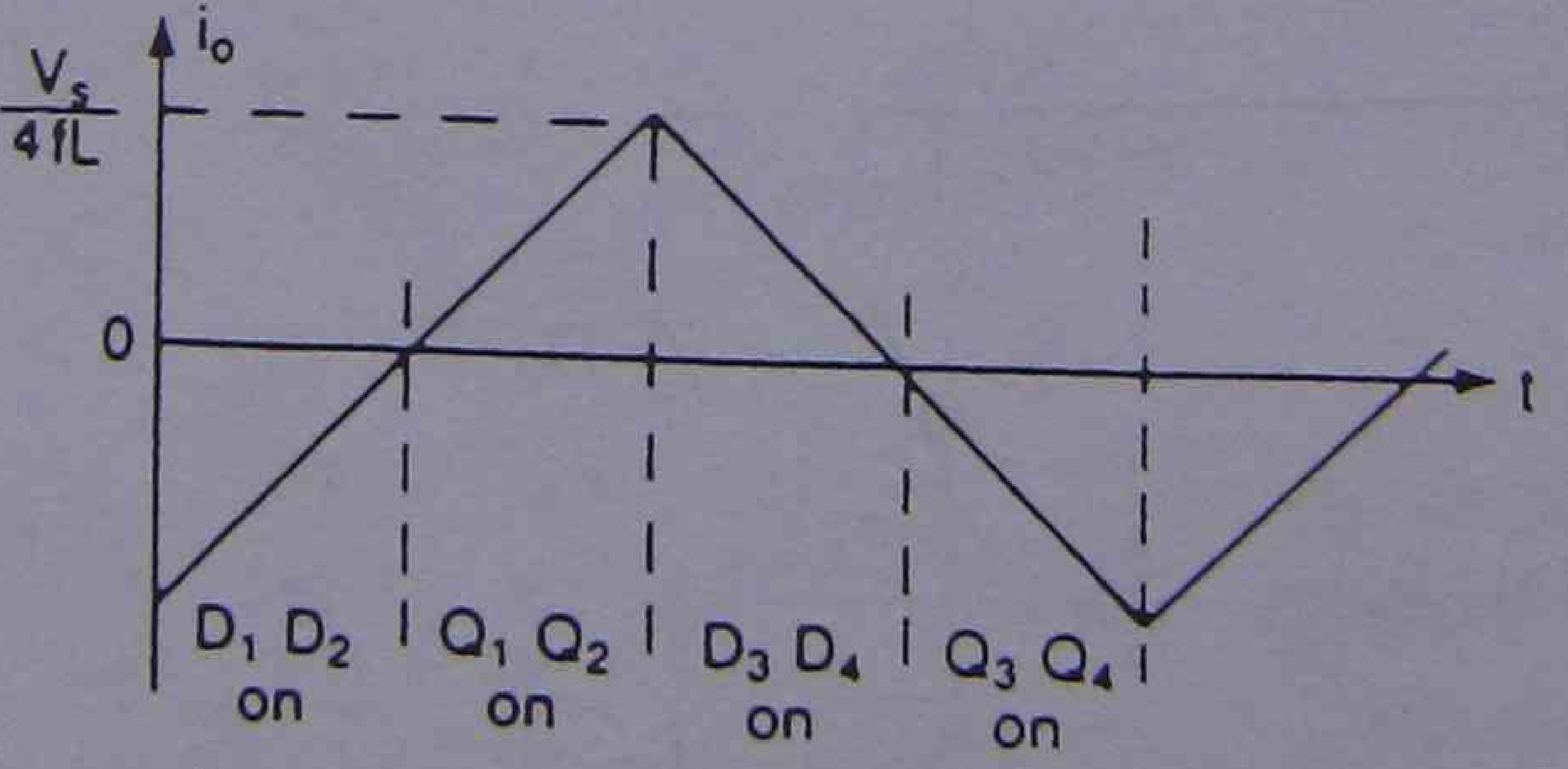
$$V_1 = \frac{4V_s}{\sqrt{2}\,\pi} = 0.90V_s$$

gives the rms value of fundamental component



(a) Circuit

(b) Waveforms



(c) Load current with highly inductive load

Figure 10-2 Single-phase full-bridge inverter.

When diodes D_1 and D_2 conduct, the energy is fed back to the dc source and they are known as *feedback diodes*. Figure 10-2c shows the waveform of load current for an inductive load.

Sydney Institute of Technology

School of Electrotechnology

Learners Theory
Work Book

Module NE76

A.C. Motor Control

Area ....

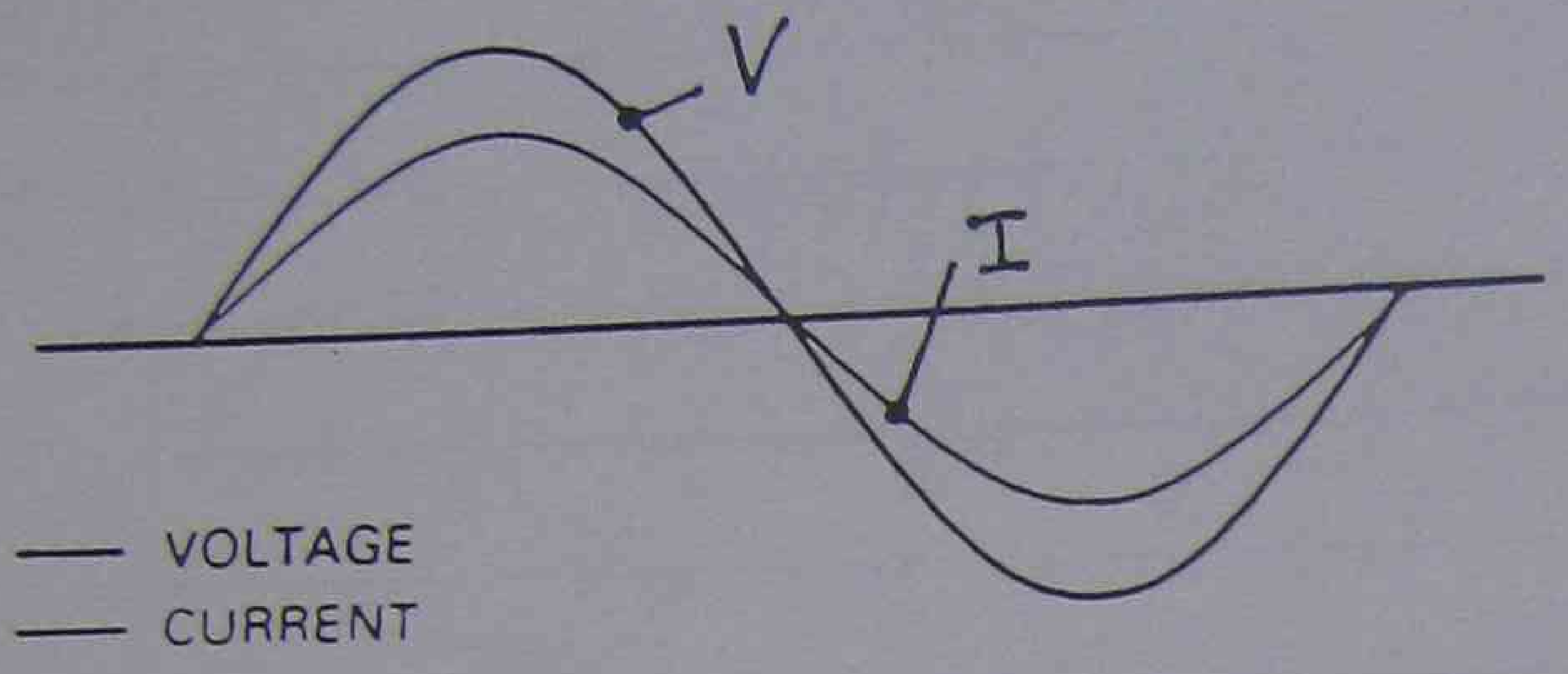
Topic ....

Session No. 5·1

For further information on this module, or this subject
contact Jim Hafford, (02) 217 3620, Bld. K. Ultimo, S.I.T.

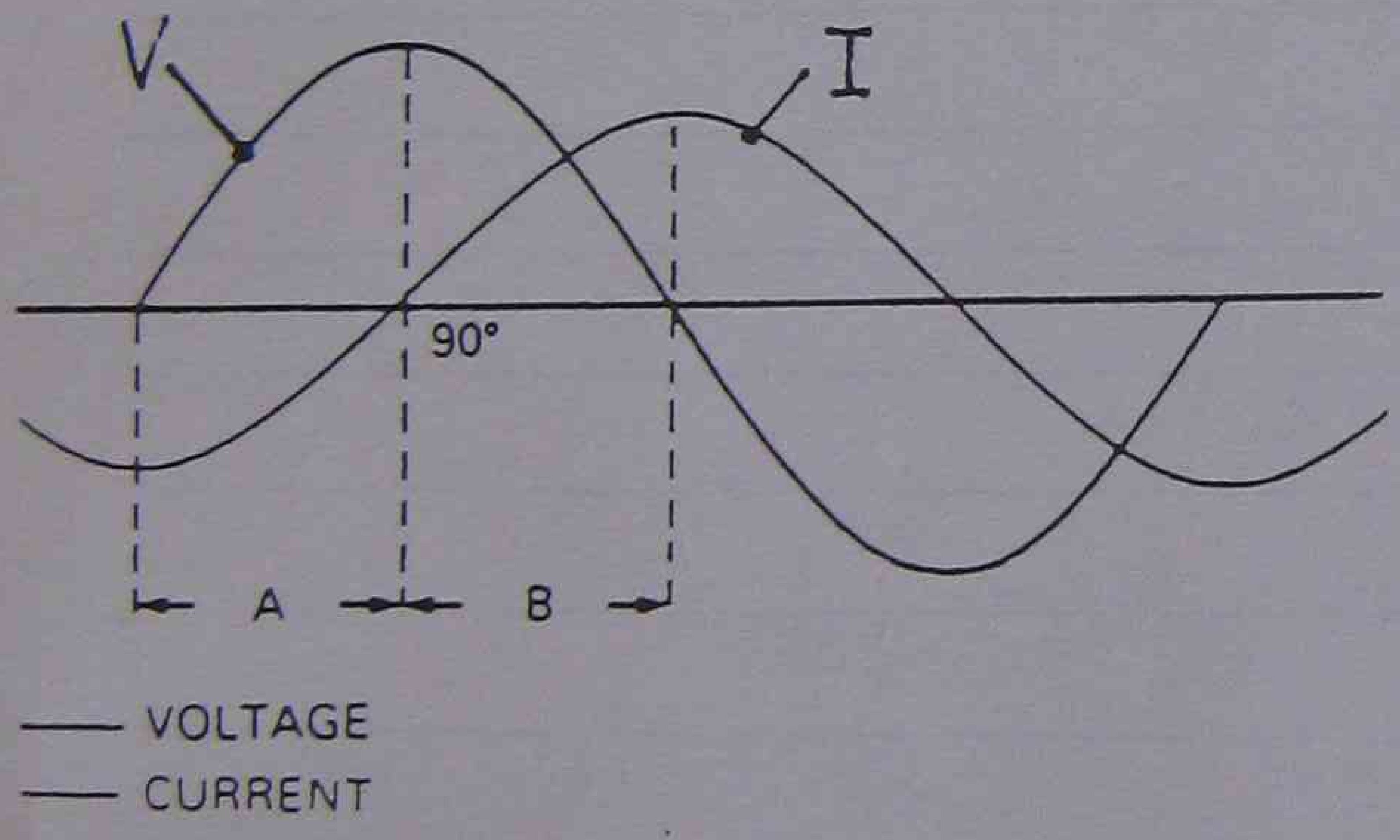
Power Factor

Typical voltage and current waveforms for a resistive load connected to an ac supply are shown in Fig.10. The two waveforms are in phase, or to express the relationship in another way, the voltage and current have the same polarity at any instant (thus the zero crossing points coincide).



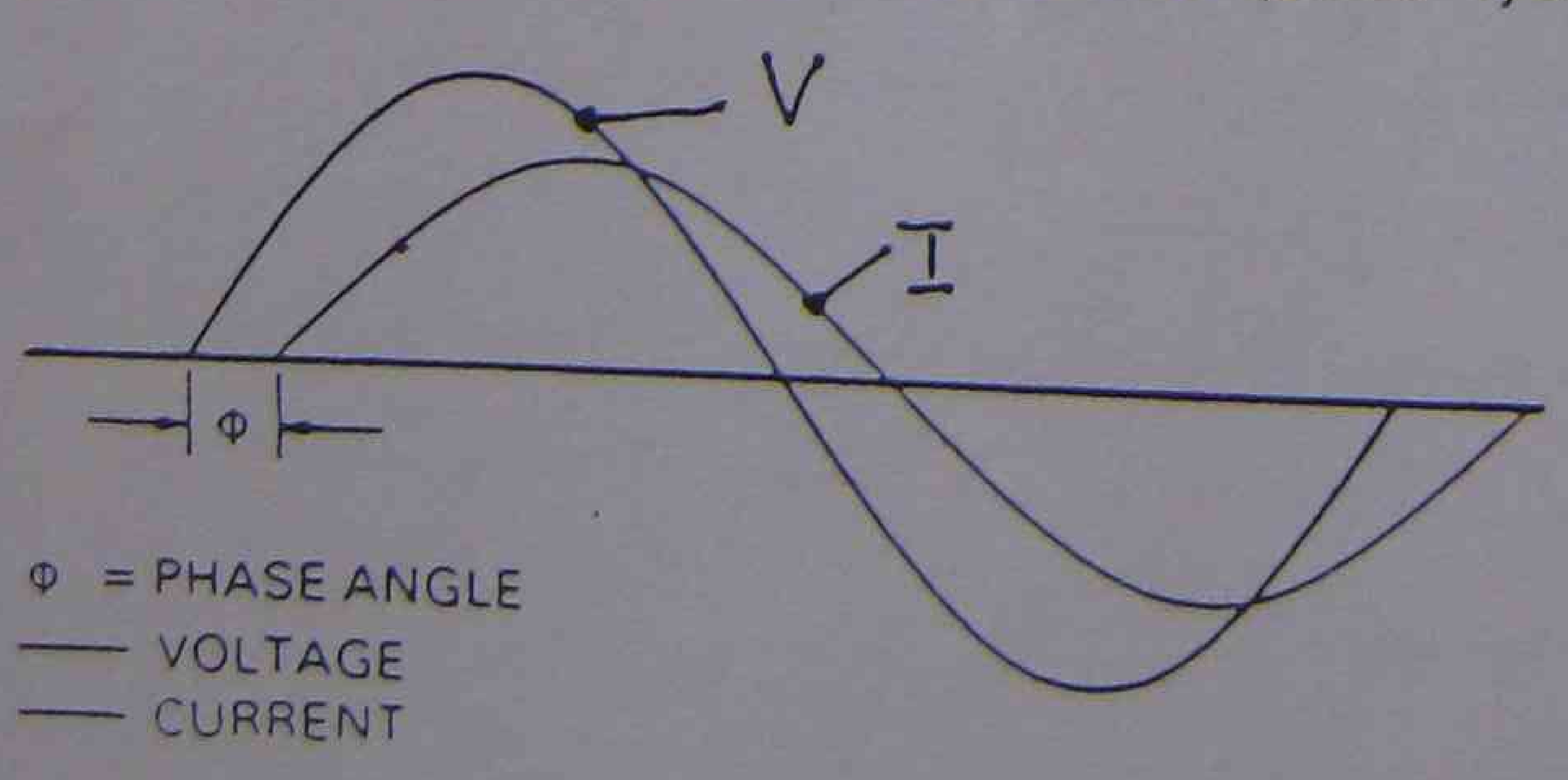
10 *Voltage and current waveforms – purely resistive load*

Since power is the product of voltage and current, it can be seen that during each half cycle the result is always positive, and therefore all the resultant power is available to be used.



11 *Voltage and current waveforms – purely inductive load*

In contrast, consider the waveforms shown in Fig.11, the load in this case being a pure inductance. The current waveform lags behind the voltage waveform by 90° due to the effect of the load inductance, and it can be seen that during the first quarter-cycle (A) the voltage and the current are of opposite polarity, whilst in the second quarter-cycle (B) they are of the same polarity. Therefore the resultant power for the half-cycle being the product of voltage and current, is zero. In this case, no useful power is produced, and the product of rms voltage and

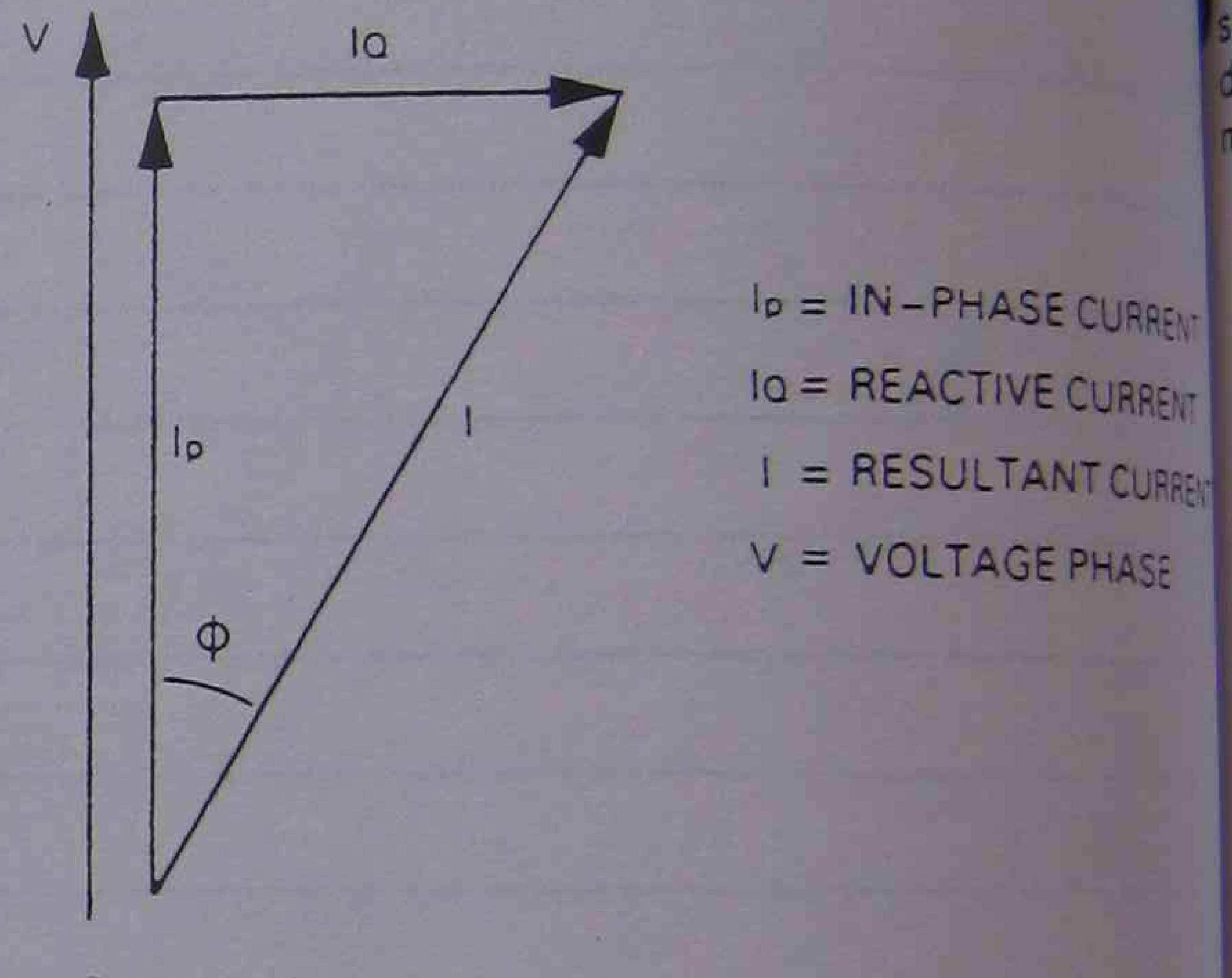


12 *Voltage and current waveforms for a load having both resistance and inductance*

In Fig.12, a load having both resistance and inductance connected to the supply and, as expected, the curr waveform lags behind the voltage waveform, the degree displacement, φ, being referred to as the phase angle this case, the current waveform has a component which is phase with the voltage waveform (due to the resist property of the load) and a component which lags voltage waveform by 90° (due to the inductive propert the load). The resulting phase angle is therefore between and 90°, and lagging, as shown in Fig.13. Clearly, the ratio in-phase current to total current is given by the cosine of phase angle, therefore:

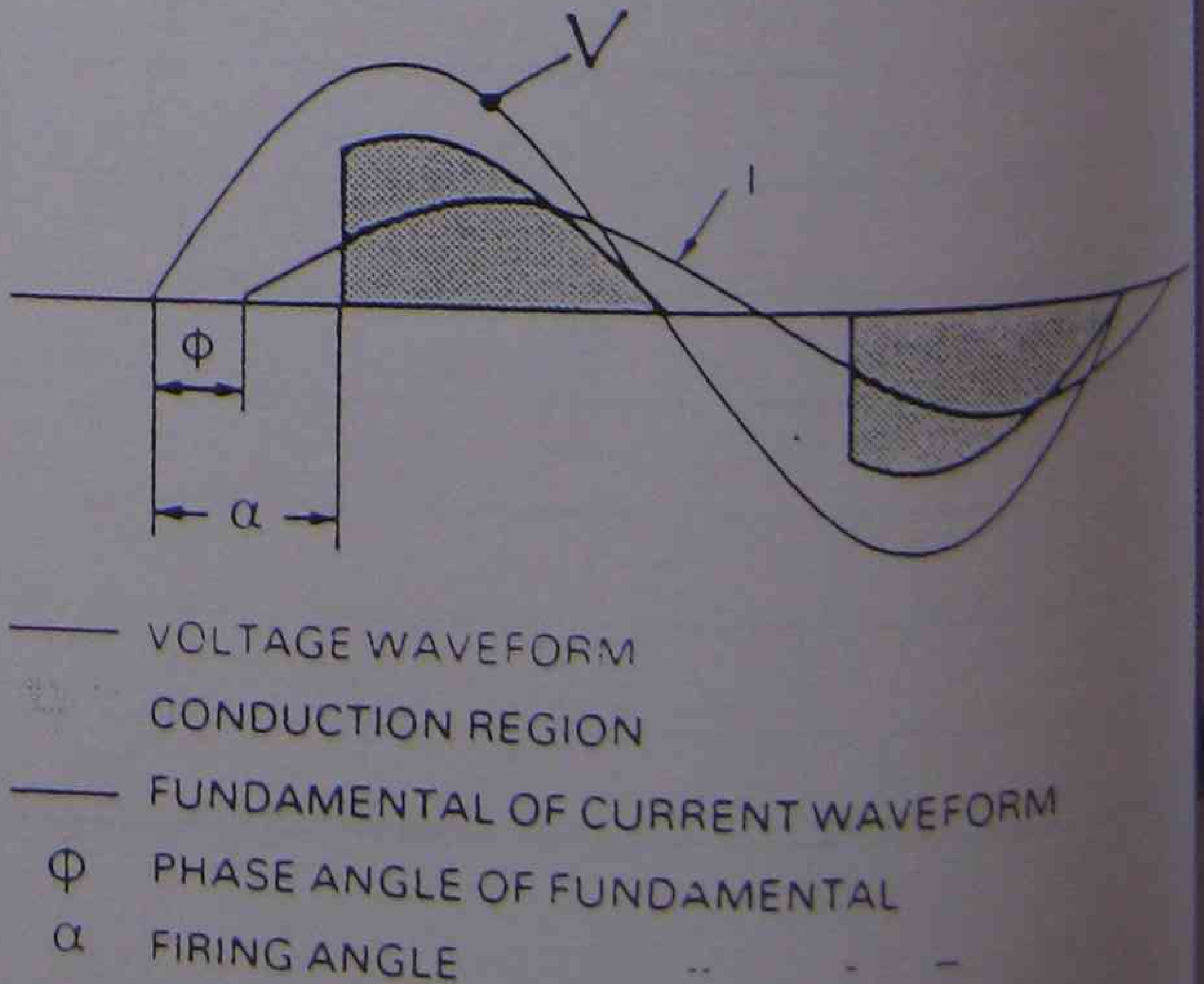
$$\cos\varphi = \frac{\text{WATTS}}{\text{VA}}.$$

Cos φ is known as the Power Factor (pf).



13 *Phasor diagram for a load having both resistance and inductance*

Voltage and current waveforms for a thyristor drive are shown in Fig.14. The shaded areas indicate the periods during which the thyristors are in conduction, having a firing angle α. The current waveform is therefore non-sinusoidal but can be resolved by Fourier analysis into a number of sinusoidal waves of different frequencies and amplitudes which, when combined, yield the actual waveform. This process results in a fundamental frequency, and harmonics which are integer multiples of the fundamental



The fundamental I_L is shown in Fig.14, and it can be seen that a phase angle φ_1 exists between it and the voltage waveform, resulting in a lagging power factor analogous to that obtained with an inductive load.

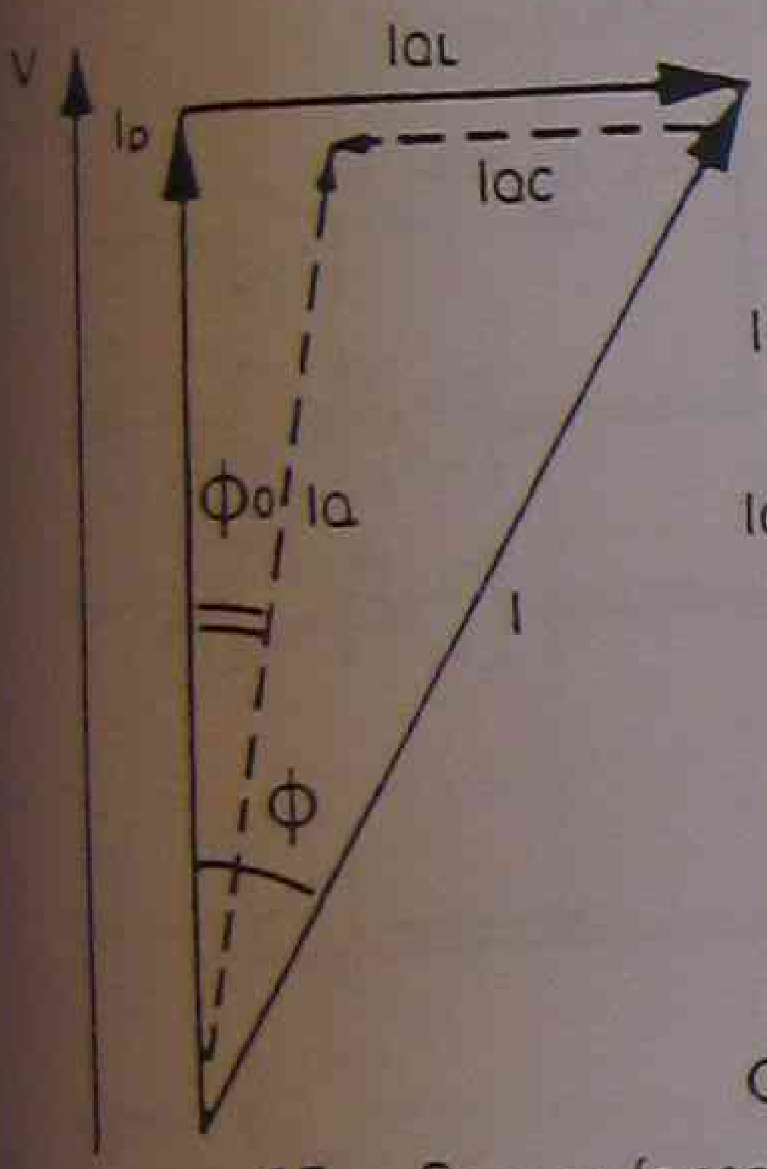
A practical way of calculating the power factor is to consider the ac in-phase component (power) to be equal to the dc output power; then the expression for power factor becomes,

$$pf = \frac{W}{VA} = \frac{V_{DC} \times I_{DC}}{\sqrt{3} \times V_{LINE} \times I_{LINE}}$$

Line current is often calculated as 0.833 x dc current, in which case

$$pf = \frac{V_{DC} \times I_{DC}}{\sqrt{3} \times V_{LINE} \times 0.833 \times I_{DC}} = \frac{0.7\,V_{DC}}{V_{LINE}}$$

This expression shows that the power factor of a thyristor drive is dependent on output voltage and, therefore, on speed. At maximum speed, the power factor of a thyristor drive is roughly comparable with that of an ac induction motor connected direct-on-line.

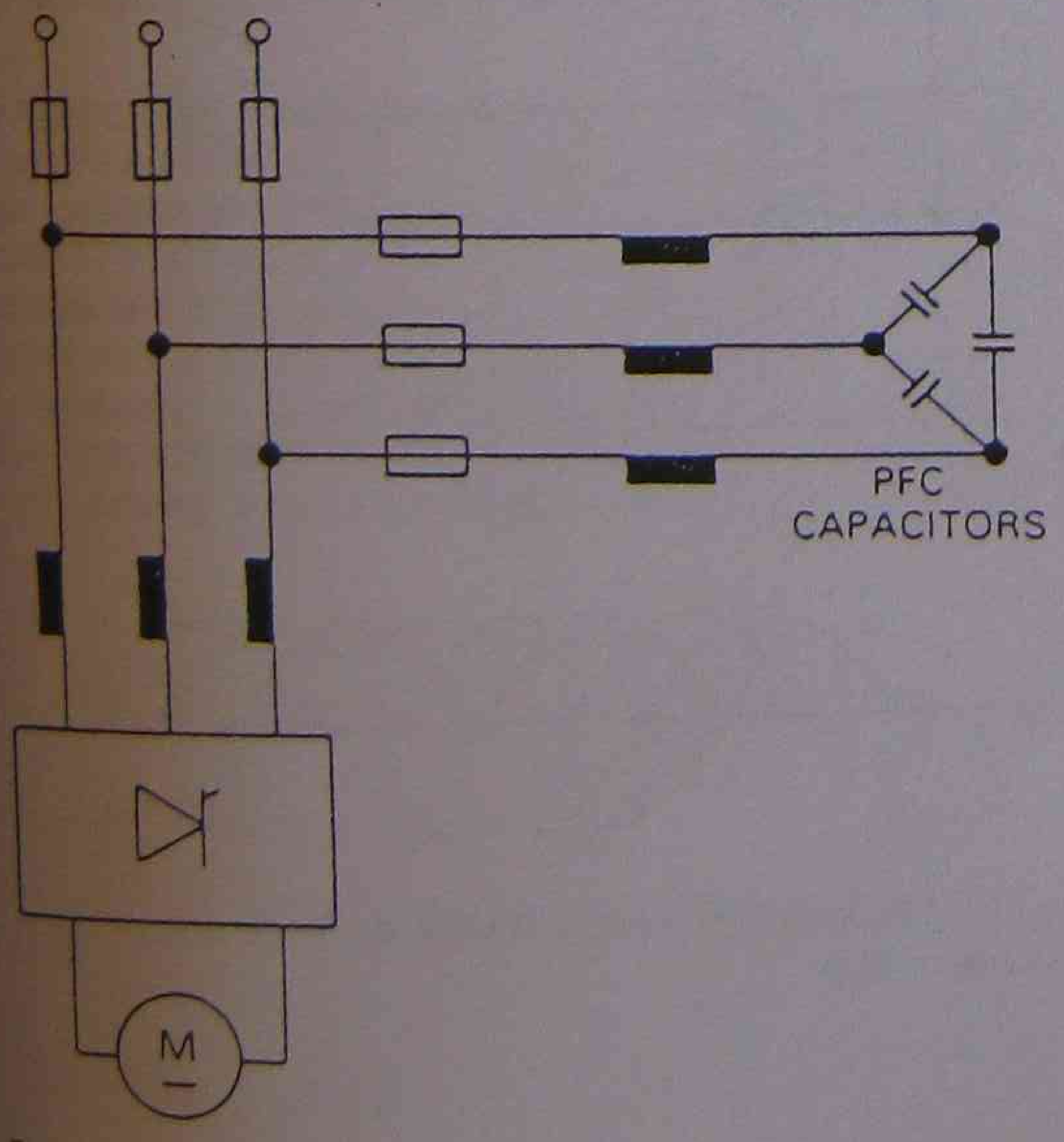


V = VOLTAGE PHASE
I_D = IN-PHASE CURRENT
I_{QL} = REACTIVE CURRENT (inductive)
I_{QC} = REACTIVE CURRENT (capacitive)
I = UNCORRECTED RESULTANT
I_Q = CORRECTED RESULTANT
Φ = UNCORRECTED PHASE ANGLE
Φ_0 = CORRECTED PHASE ANGLE

15 *Power factor correction*

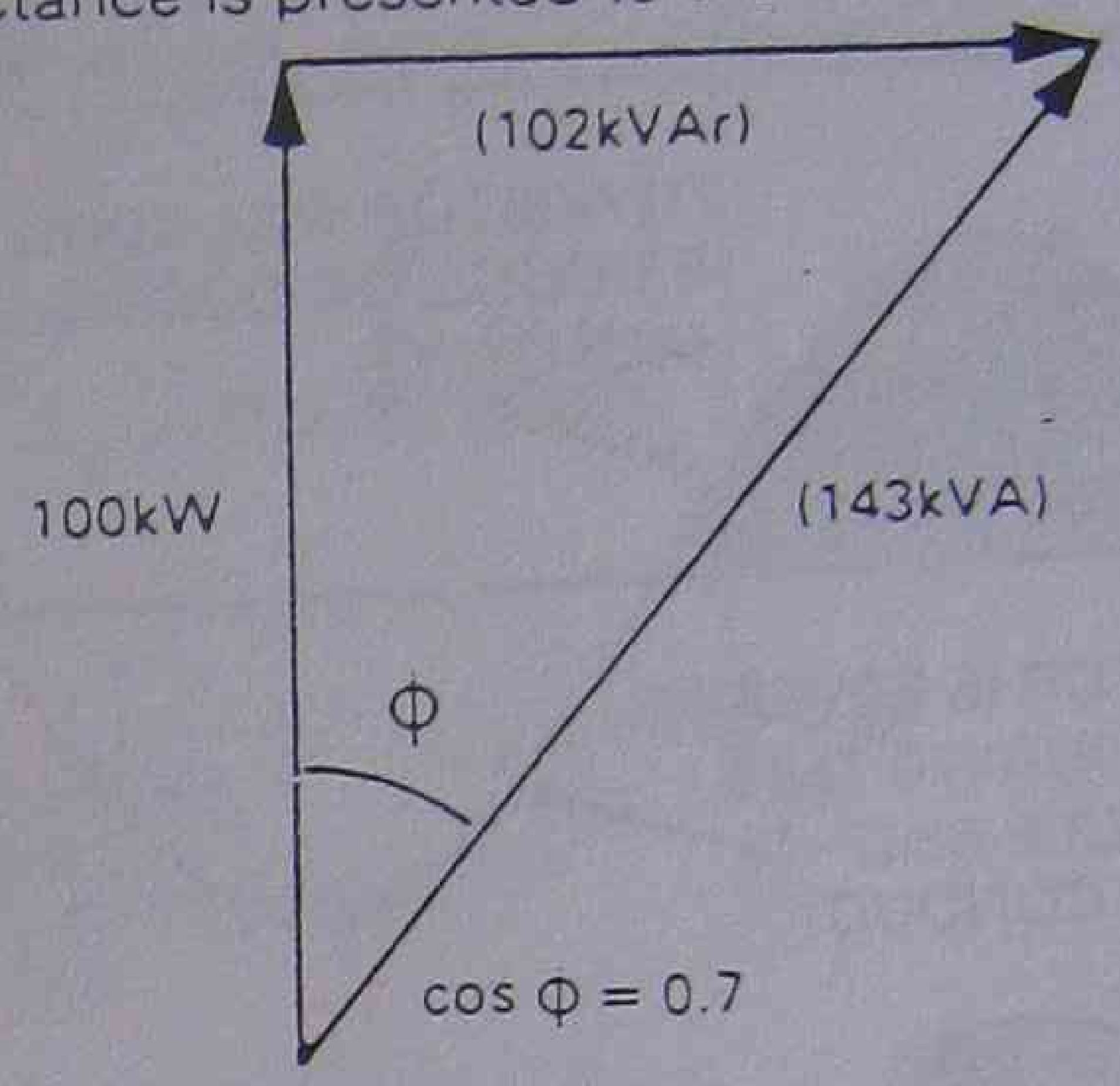
With larger drives it is often desirable to employ power factor correction (PFC) in order to improve the power factor of the total load connected to the supply.

This is done by means of capacitors connected between the supply lines, and chosen to give the required correction at a particular speed, typically correcting to a pf of 0.9 at full speed, although capacitors can be switched to accommodate other speeds should the application demand it. Fig.15 illustrates the principle, which is based on the fact that a capacitive load results in a leading phase angle, which can be used to offset the lagging phase angle of the thyristor drive.



16 *Typical circuit for power factor correction*

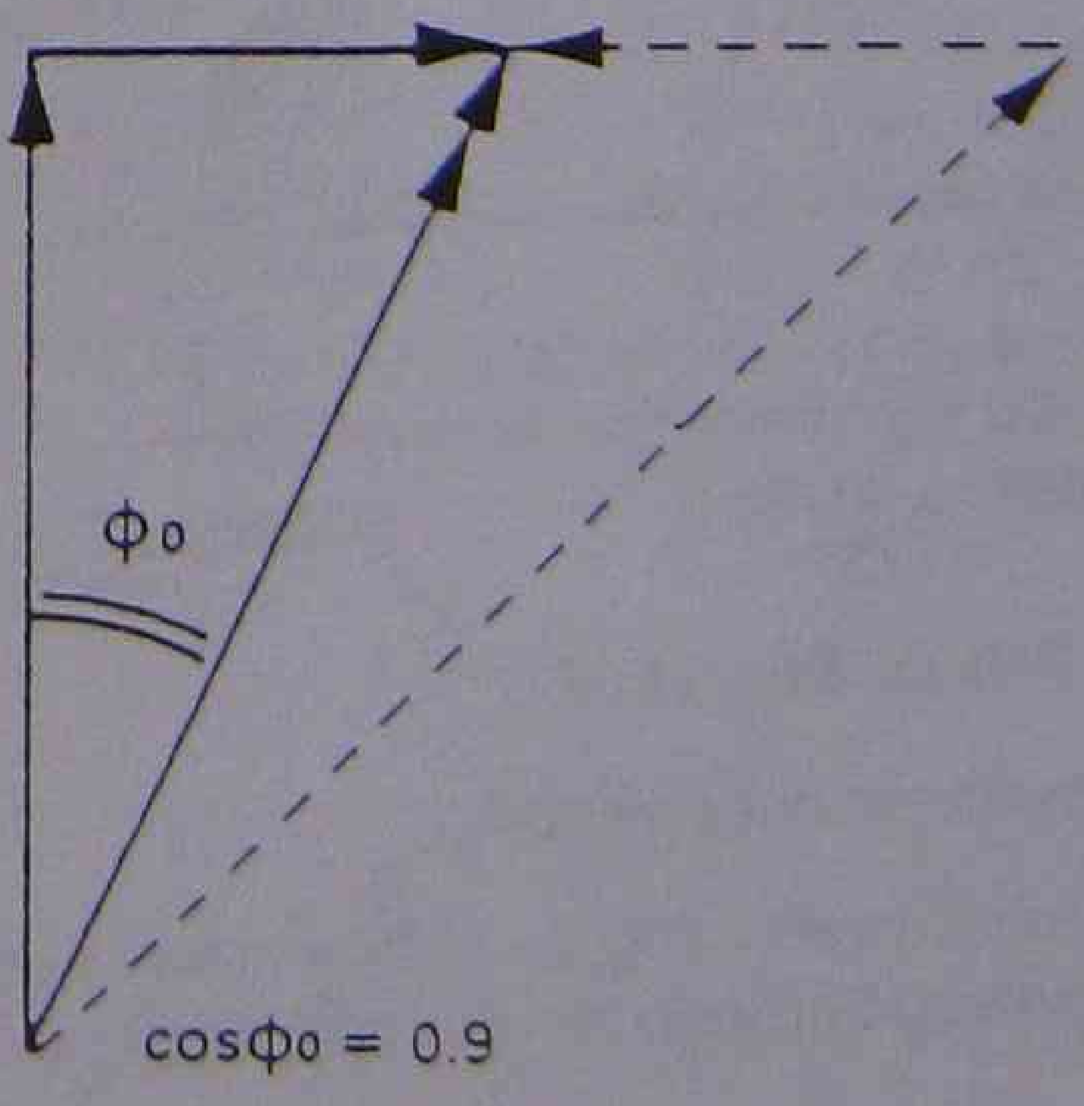
Because the capacitors present a low impedance to the high-frequency harmonics it is necessary to insert chokes between the lines and the capacitors, Fig.16, to limit harmonic currents. The resulting combination of capacitance and inductance is tuned so that a capacitive reactance is presented to the fundamental (supply) frequency, whilst an inductive reactance is presented to the harmonics.



17 *Phasor diagram for 100kW drive at pf = 0.7*

Example To calculate the PFC capacitors required to correct a 100kW drive from a pf of 0.7 to 0.9. Fig.17 shows the phase diagram of the uncorrected arrangement.

Knowing power and pf (cos φ), the kVA can be calculated:



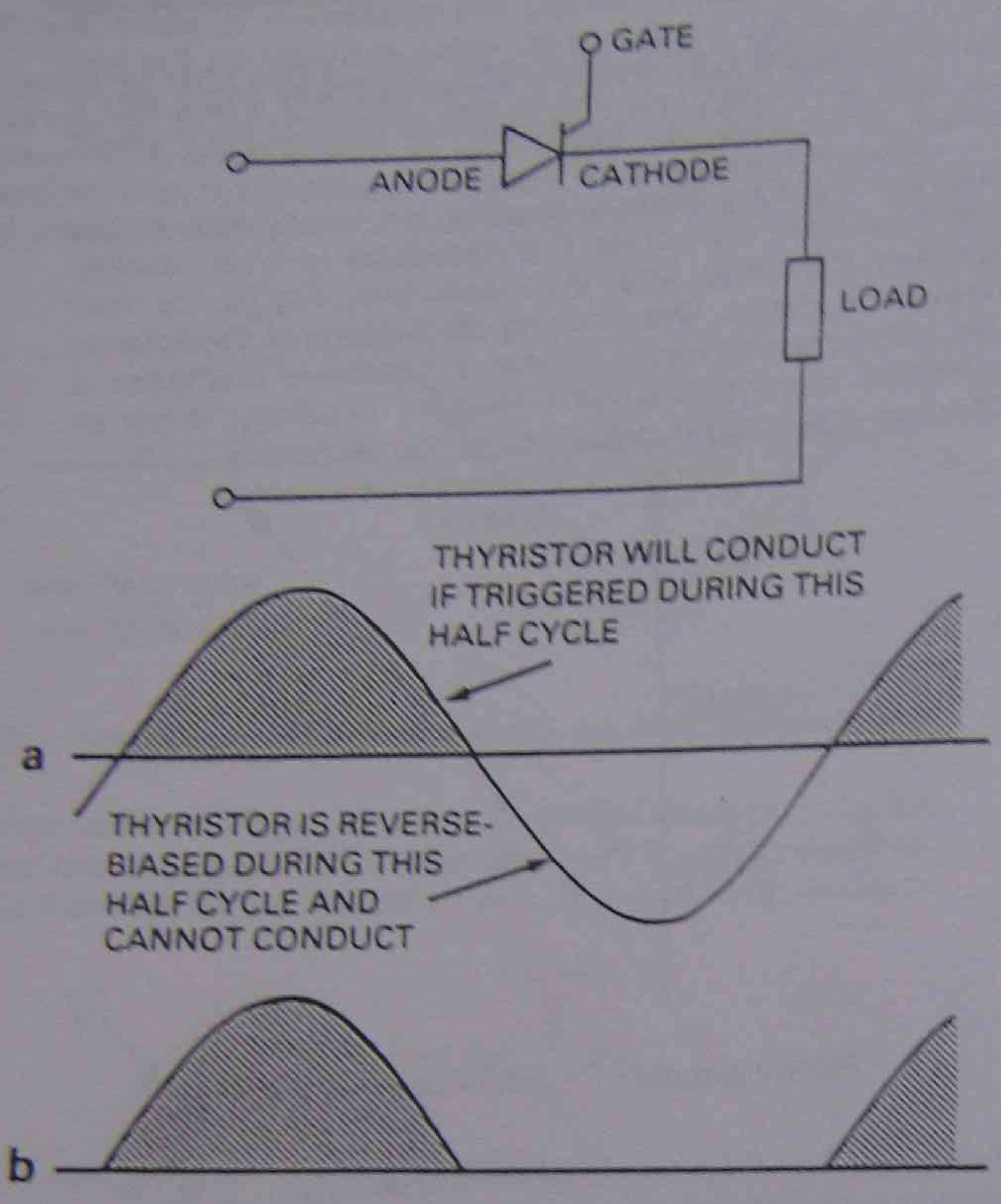
18 *Phasor diagram for 100kW drive with PF corrected to 0.9*

The capacitance required is the difference between the actual and the required reactive components:

102 − 48 = 54kVAr capacitative, required for correction, illustrated in Fig.18.

A.C. Motor Control.

Topic -

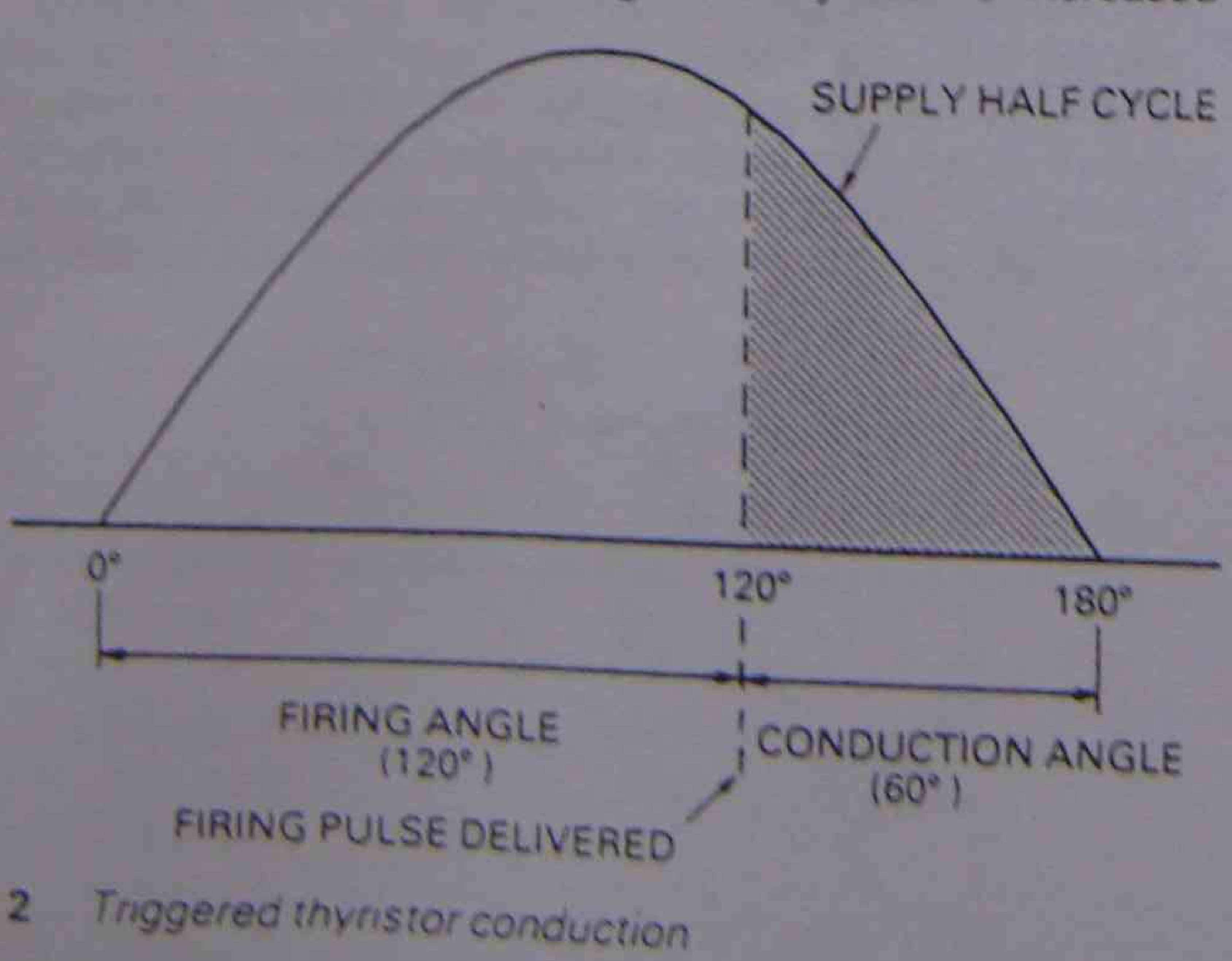


OUTPUT WAVEFORM

1 *Thyristor full wave conduction* **a** *Supply* **b** *Output*

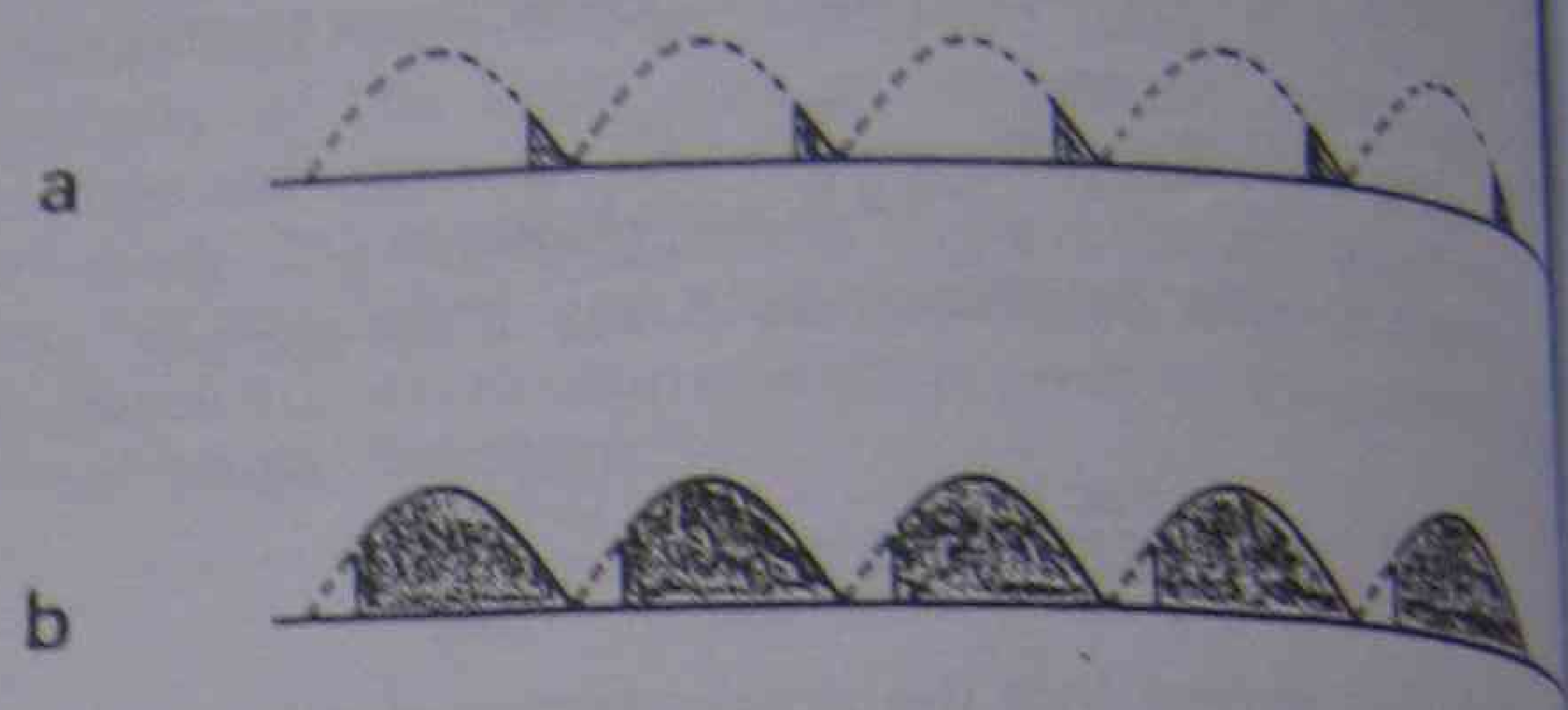
which the thyristor is fired is known as the firing angle, or phase angle, and the portion of the half-cycle during which the device is conducting is called the conduction angle, Fig.2.

It can be seen that by 'phasing forward' or advancing the firing angle from 180° to zero, to use the conventional expression, the output through the thyristor is increased from zero to maximum, Fig.3, since this progressively increases the corresponding conduction angle from zero to 180°.

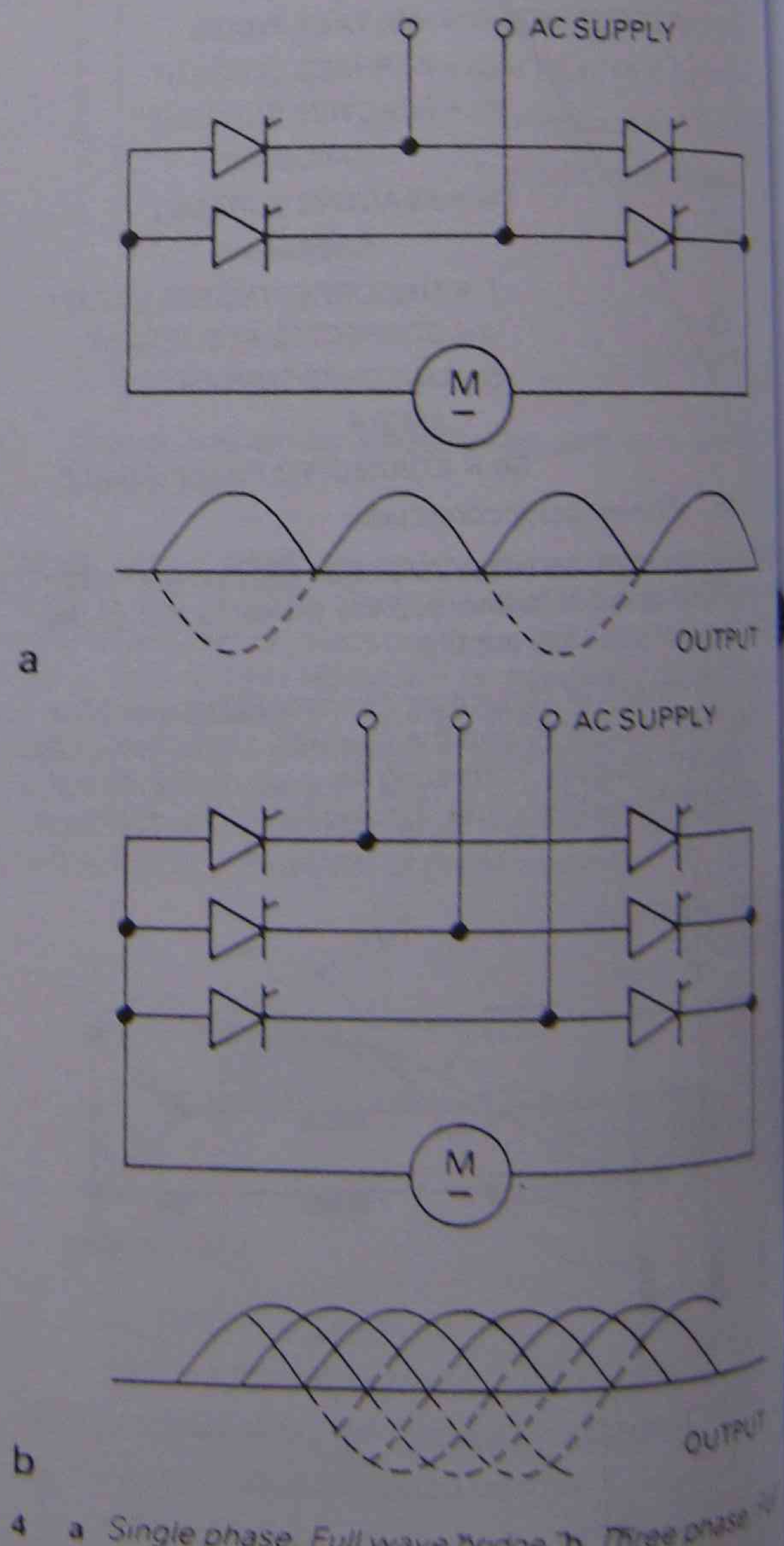


2 *Triggered thyristor conduction*

By arranging thyristors in a bridge configuration, a full-wave rectified output is obtained, which is smoother than the half-wave output of a single device, Fig.4a, and if the supply is 3-phase, the output (at 180° firing angle) is smoother still, Fig.4b.



3 *Typical output of a single-phase thyristor bridge for la and small firing angles:* **a** *Output waveform with smal conduction angle (firing pulses retarded or" phased ba* **b** *Output waveform with large conduction angle (firing pulses advanced or" phased forward").*



4 **a** *Single phase, Full wave bridge* **b** *Three phase*

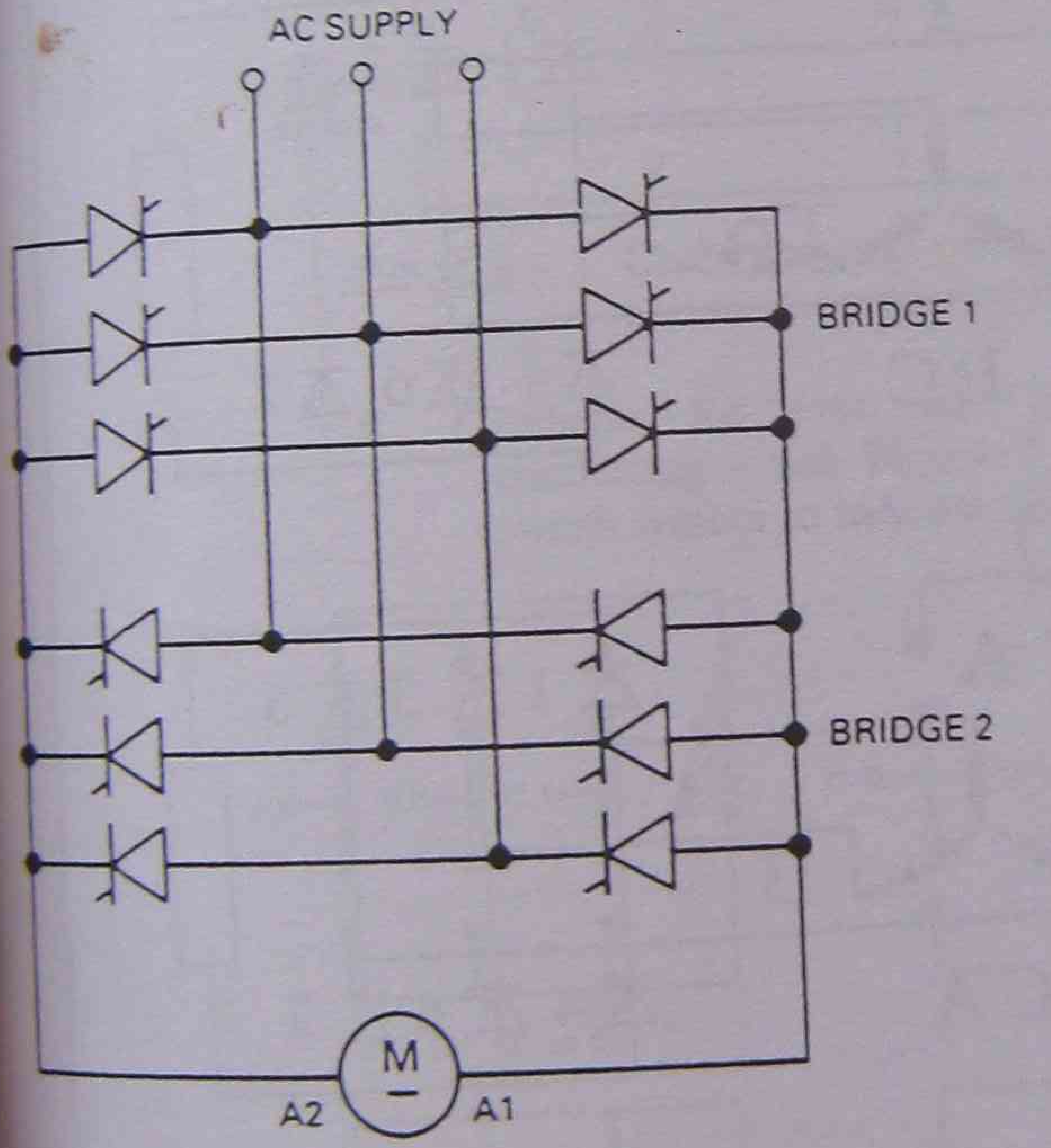
A.C. Motor Control.

Topic -

Four Quadrant Operation

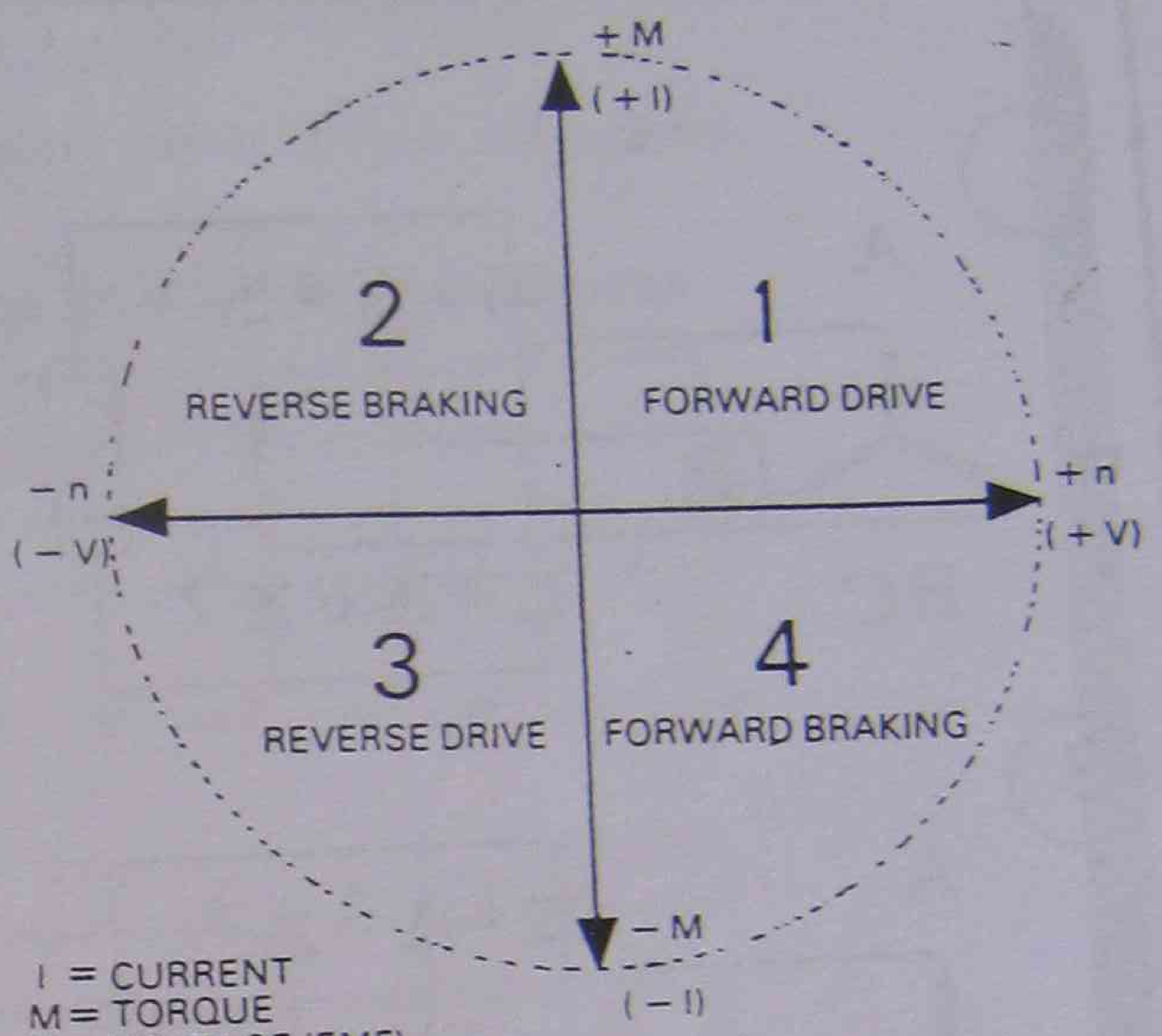
By definition, the output current of a rectifier is unidirectional, therefore the torque produced by the motor will also be in one direction or sense of rotation. In applications where only one direction of rotation is required, and the load can be allowed to coast to rest, this is adequate. However, there are many applications in which reversal of rotation is necessary, or where large load inertias must be decelerated.

In such cases, two thyristor bridges are connected in inverse-parallel as shown in Fig.5. When Bridge 1 is conducting, current flow in the motor is from A1 to A2. Current flow is reversed when Bridge 2 conducts, and therefore the motor torque reverses. This effect can be used to reverse the direction of rotation of the motor, and to provide braking torque.



5 *Power circuit of a four quadrant thyristor drive.*

The arrangement described above is known as a Four-Quadrant (4Q) drive, since it is capable of operation in any of the four quadrants of the torque-speed diagram (Fig.6). In Quadrant 1, both voltage and current, and therefore speed and torque are in the positive, or forward, direction. This is consistent with a motor driving a load, taking power from the mains. Similarly, in Quadrant 3, both speed and torque are negative; this corresponds to a motor turning in the reverse direction, driving a load and again taking power from the mains.



I = CURRENT
M = TORQUE
V = VOLTAGE (EMF)
n = SPEED

6 *The four quadrants of the dc torque speed diagram*

In quadrants 2 and 4, the speed and torque are in mutually-opposed directions, that is to say the torque of the motor is opposing its rotation, giving a braking effect. It follows, then, that the mechanical kinetic energy of the load is being converted into electrical energy; the motor is behaving as a generator and the system as a whole is delivering power into the mains. This behaviour is known as regeneration, and has two main applications. The most usual is regenerative braking of a rotating mass (eg the spindle of a machine tool or a coil of material in a process line) to give a fast, controlled stop.

Another important application of regeneration is in dynamometers, where a regenerative drive is used to provide a load for a mechanical power source (eg a diesel engine) both for testing it under load and for measuring its output.

A further advantage of the four-quadrant drive is the ease with which a motor can be reversed, simply by changing bridges. A 'single-ended' drive (having only one thyristor bridge) can only be reversed by means of contactors which reverse the connections to the motor armature or field (Fig.7). The contactors must be interlocked so that they cannot operate together and cause a short-circuit. The arrangement obviously introduces an operational delay and the need for external control circuits. A 4Q drive, on the other hand, can be quickly and simply reversed by switching from the firing circuit of one bridge to the other.

A.C. Motor Control.

Topic - ____________________

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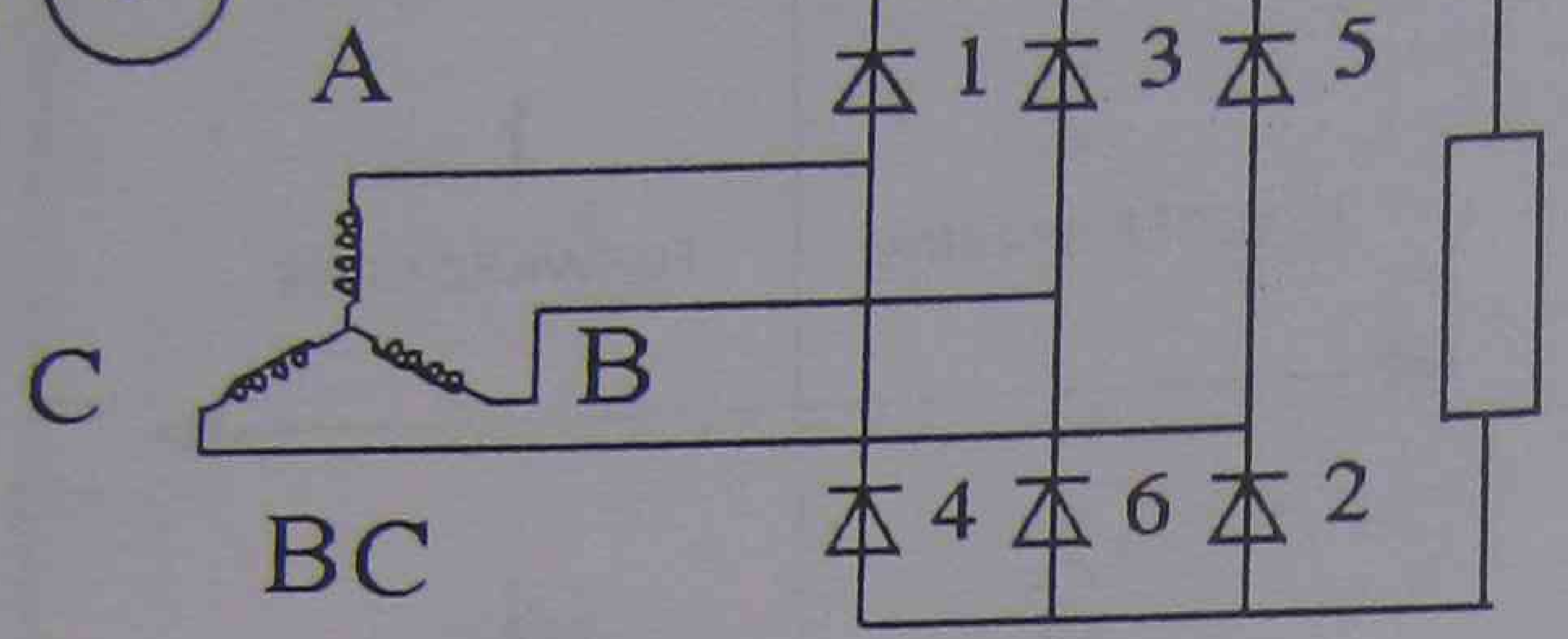

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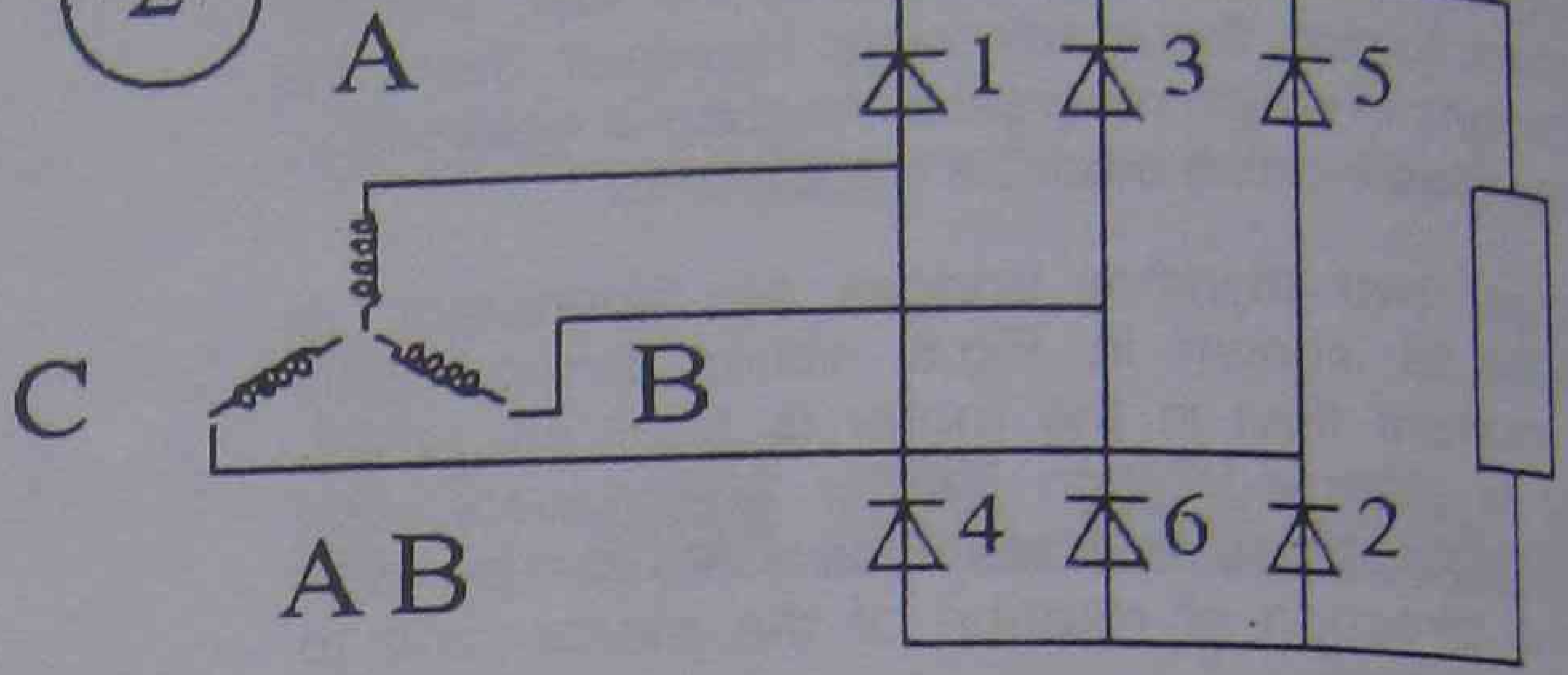

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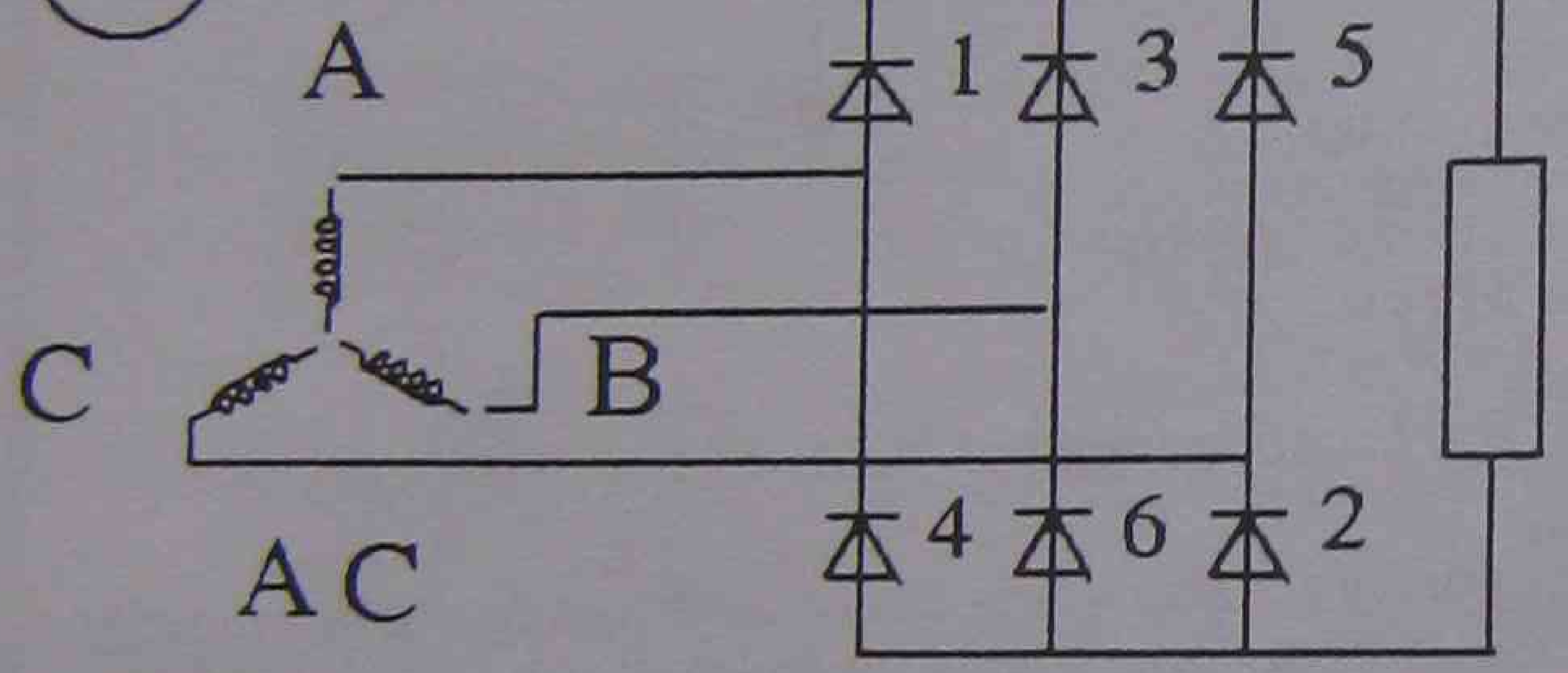

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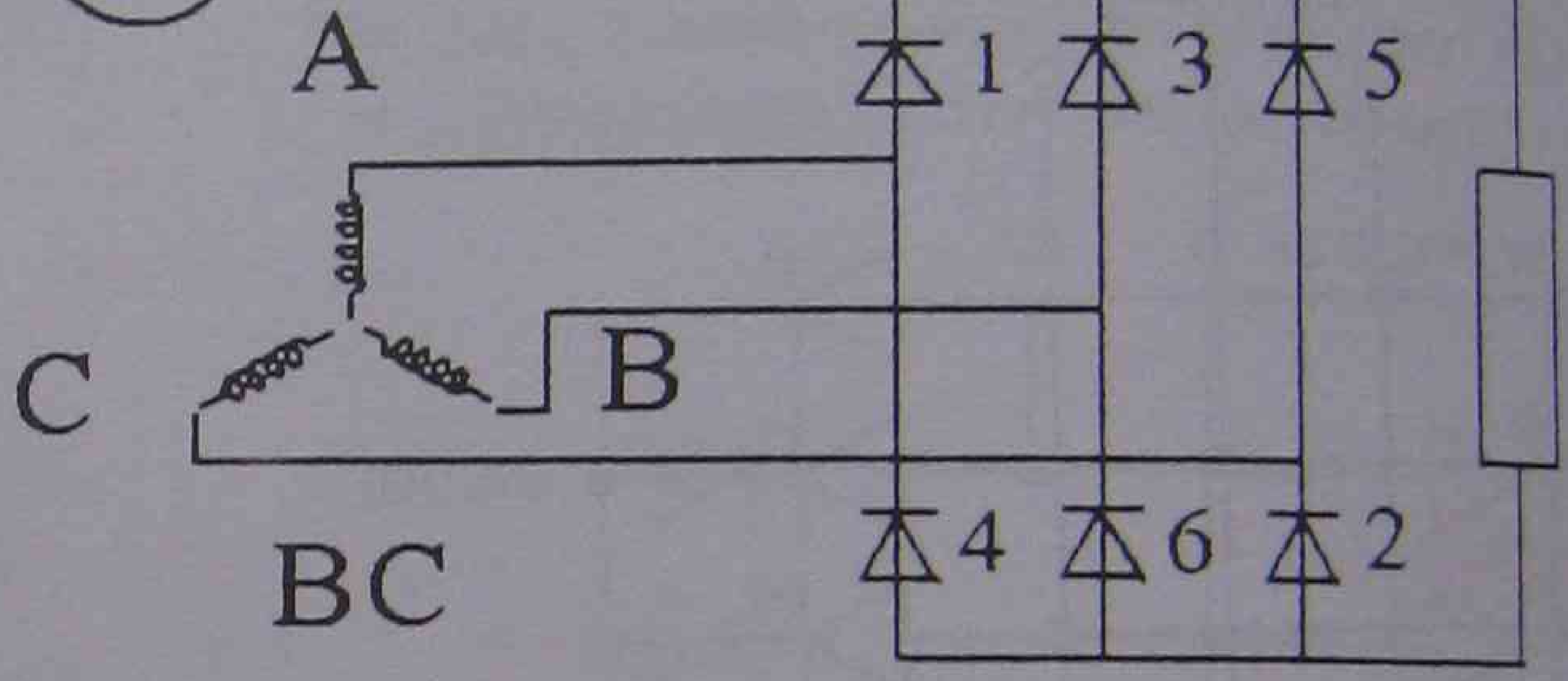

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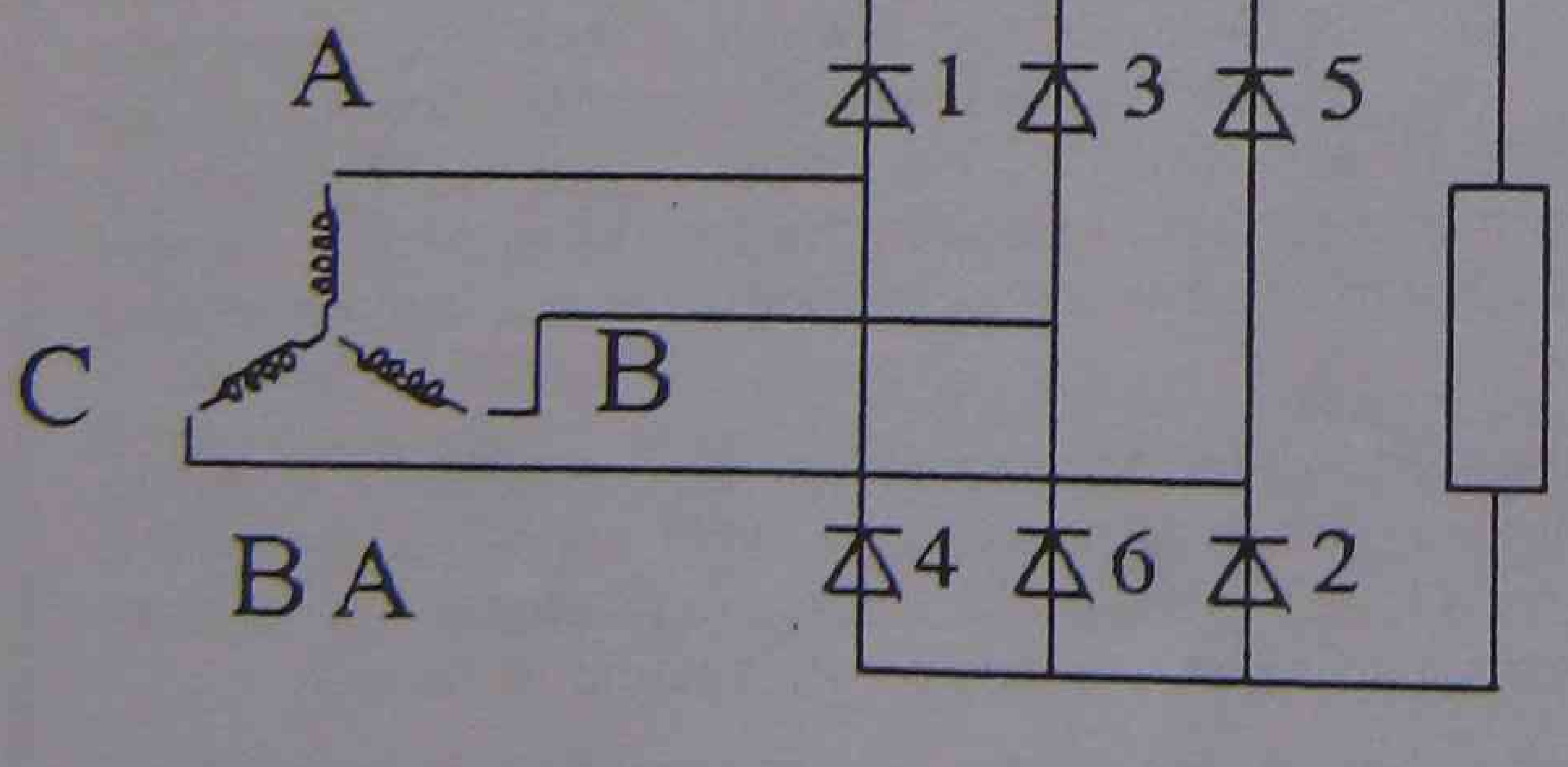

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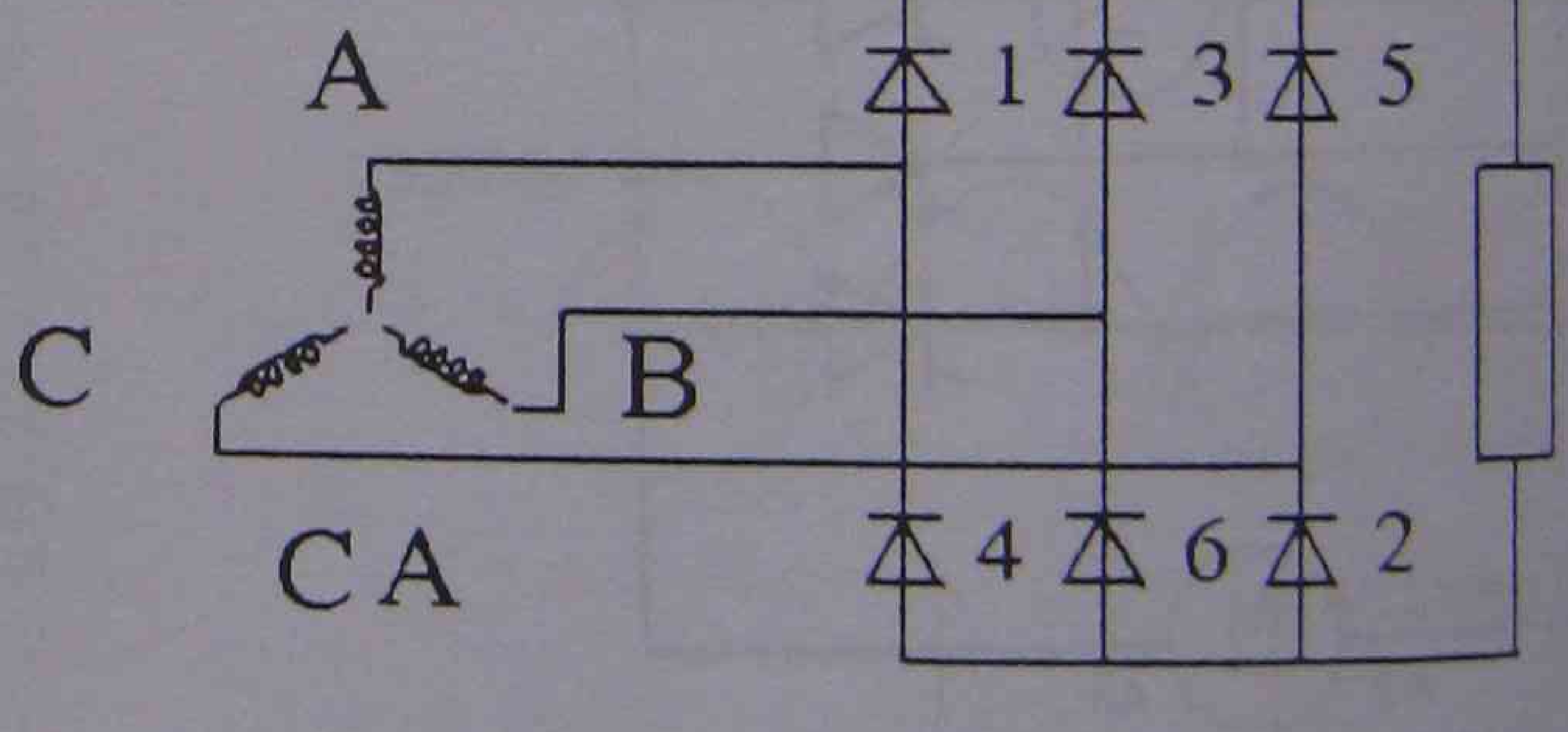
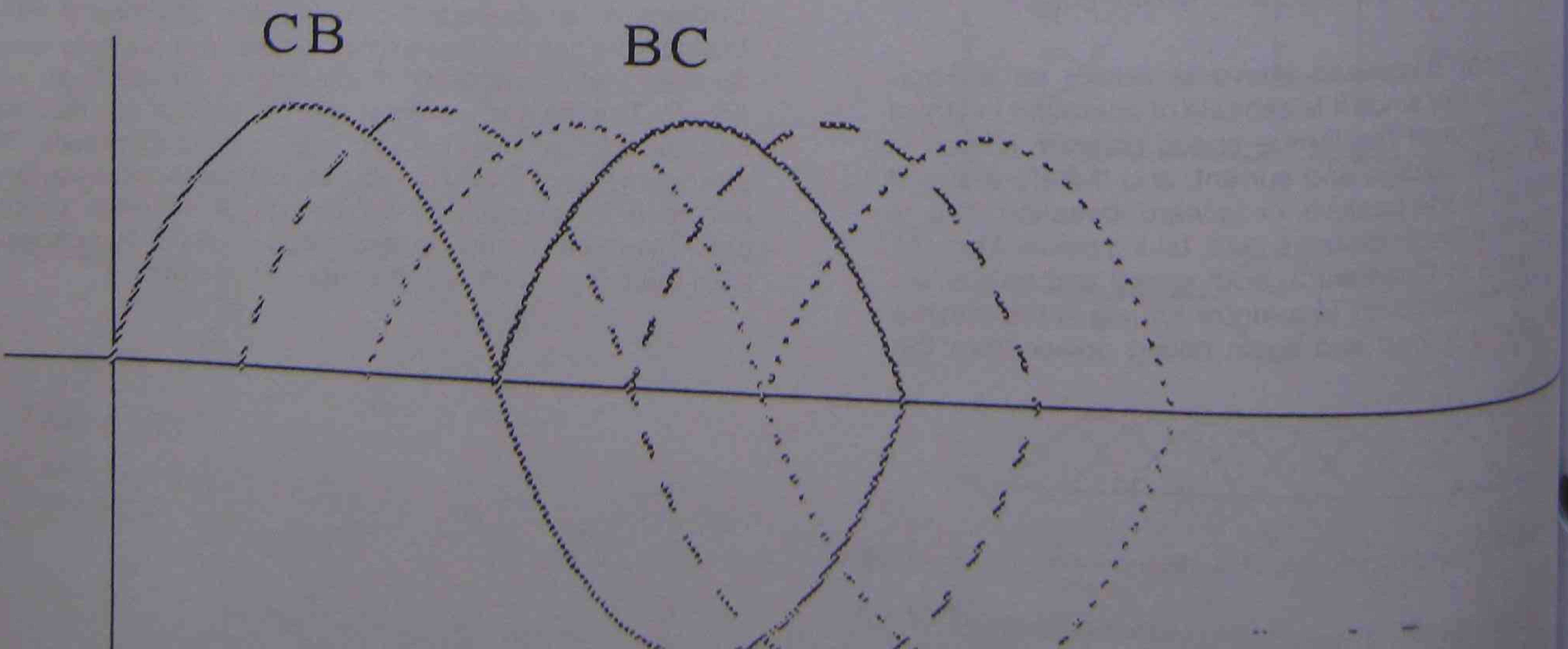




(c) Operation of Three Phase Full Wave Rectifier

- List inside each rectangle, the highest positive voltage and negative voltage for the periods indicated.
- List which diode would be conducting (ON) for each period indicated

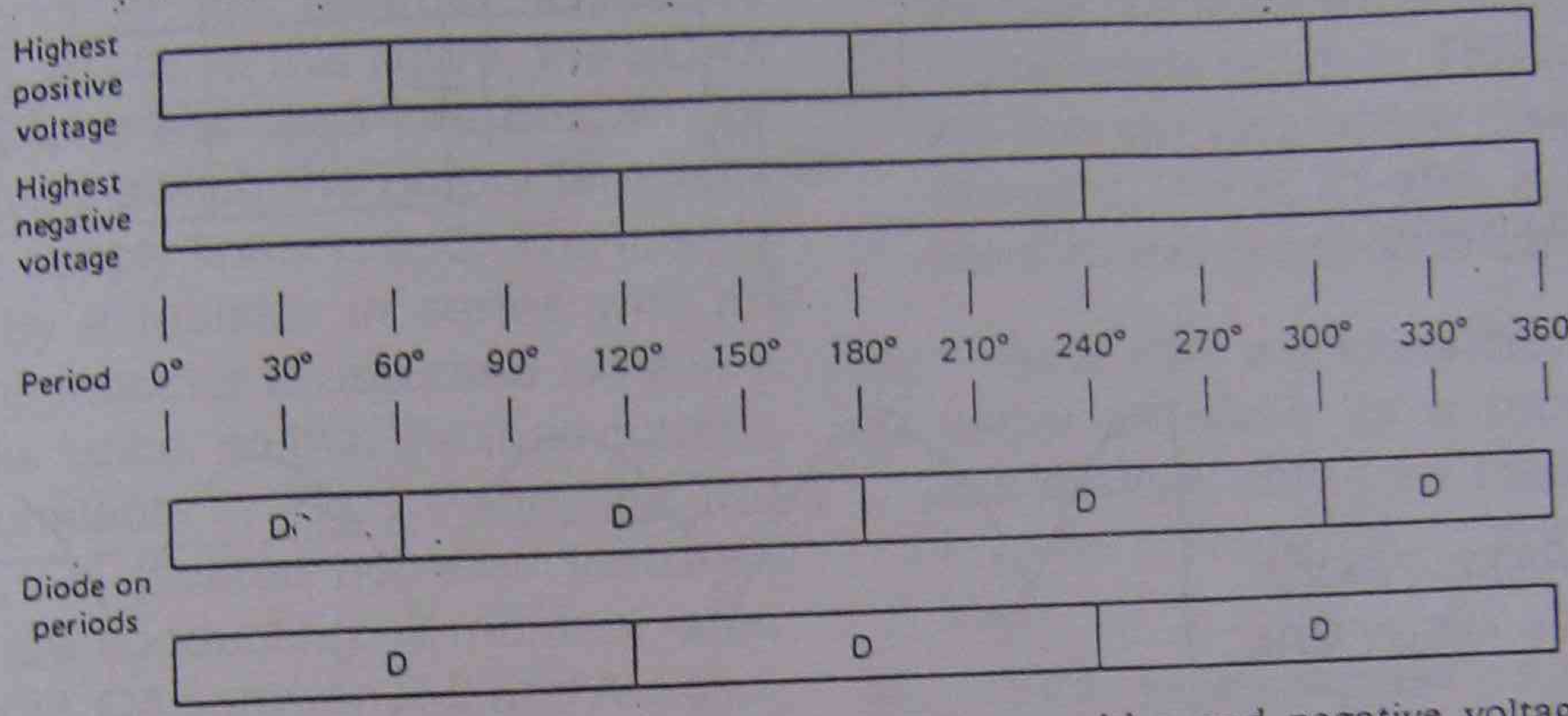
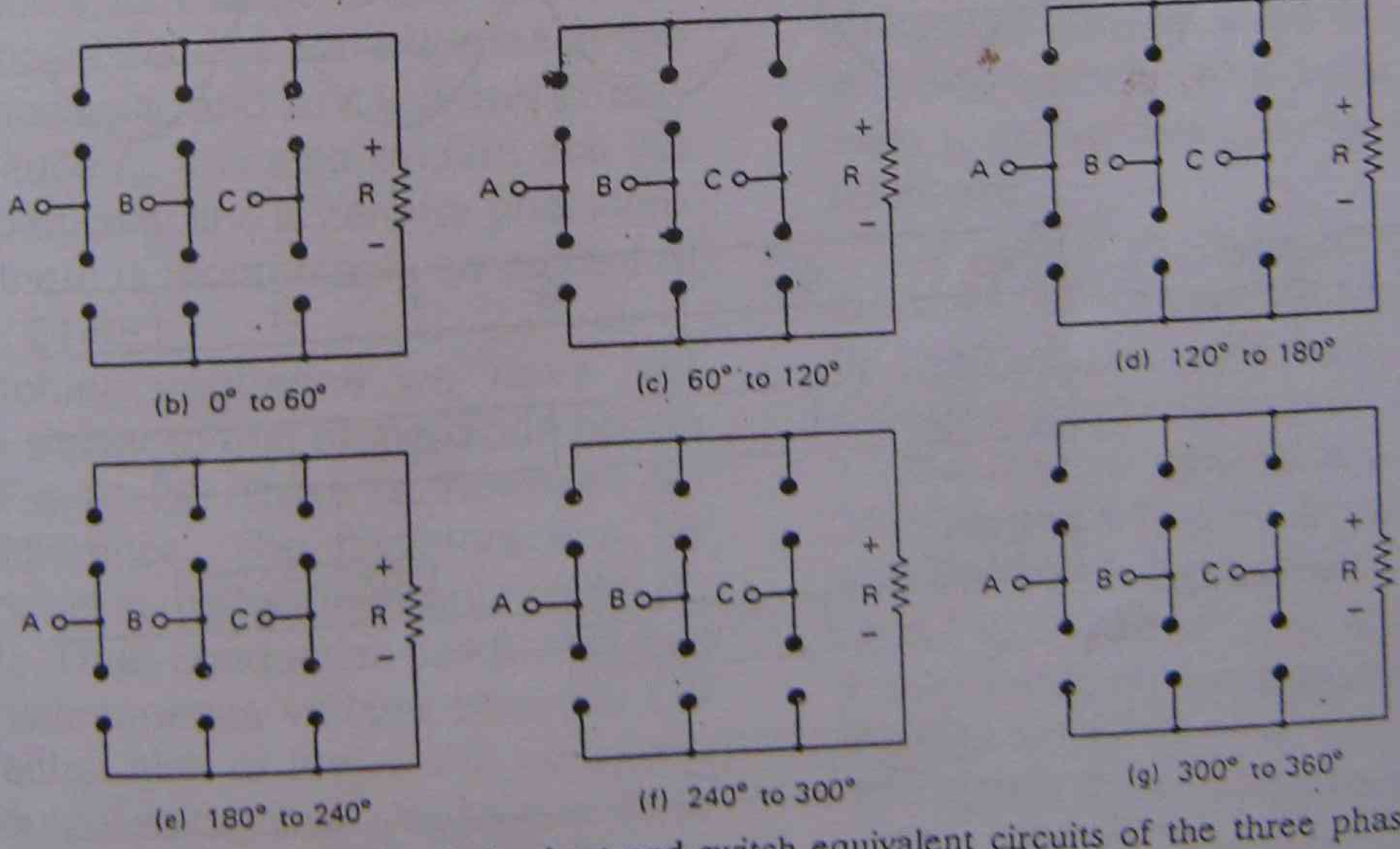


Figure 8 - Chart showing periods of highest positive and negative voltages and their respective ON diodes.

- Indicate for each 60° period which two (2) diodes would be conducting - use Figure 8 results above and use the standard close switch symbol to indicate the ON diode.



(b) 0° to 60°
(c) 60° to 120°
(d) 120° to 180°
(e) 180° to 240°
(f) 240° to 300°
(g) 300° to 360°

Figure 9 - ON diode chart and switch equivalent circuits of the three phase full wave rectifier.

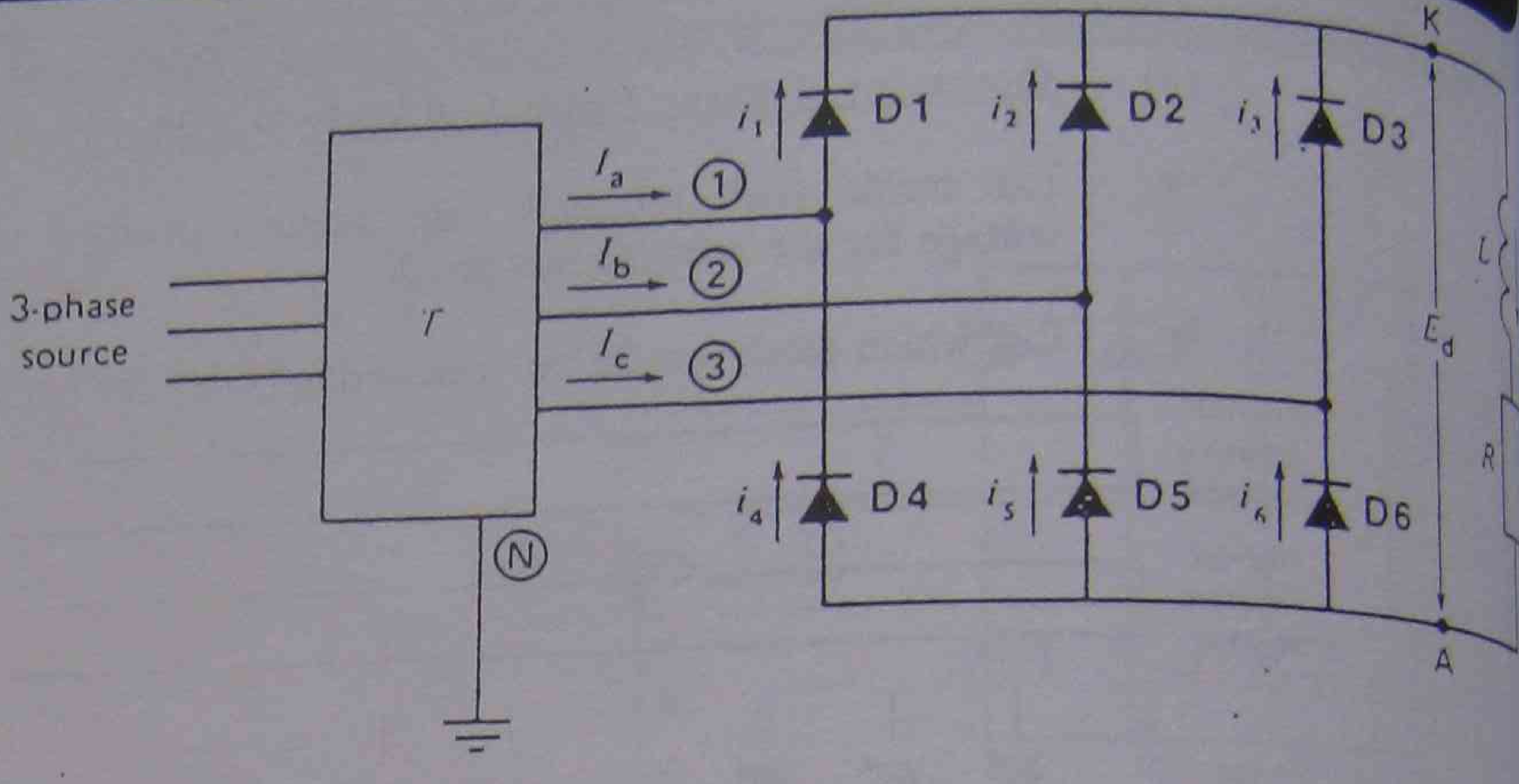


Figure 21-19
Three-phase, 6-pulse rectifier
with inductive filter.

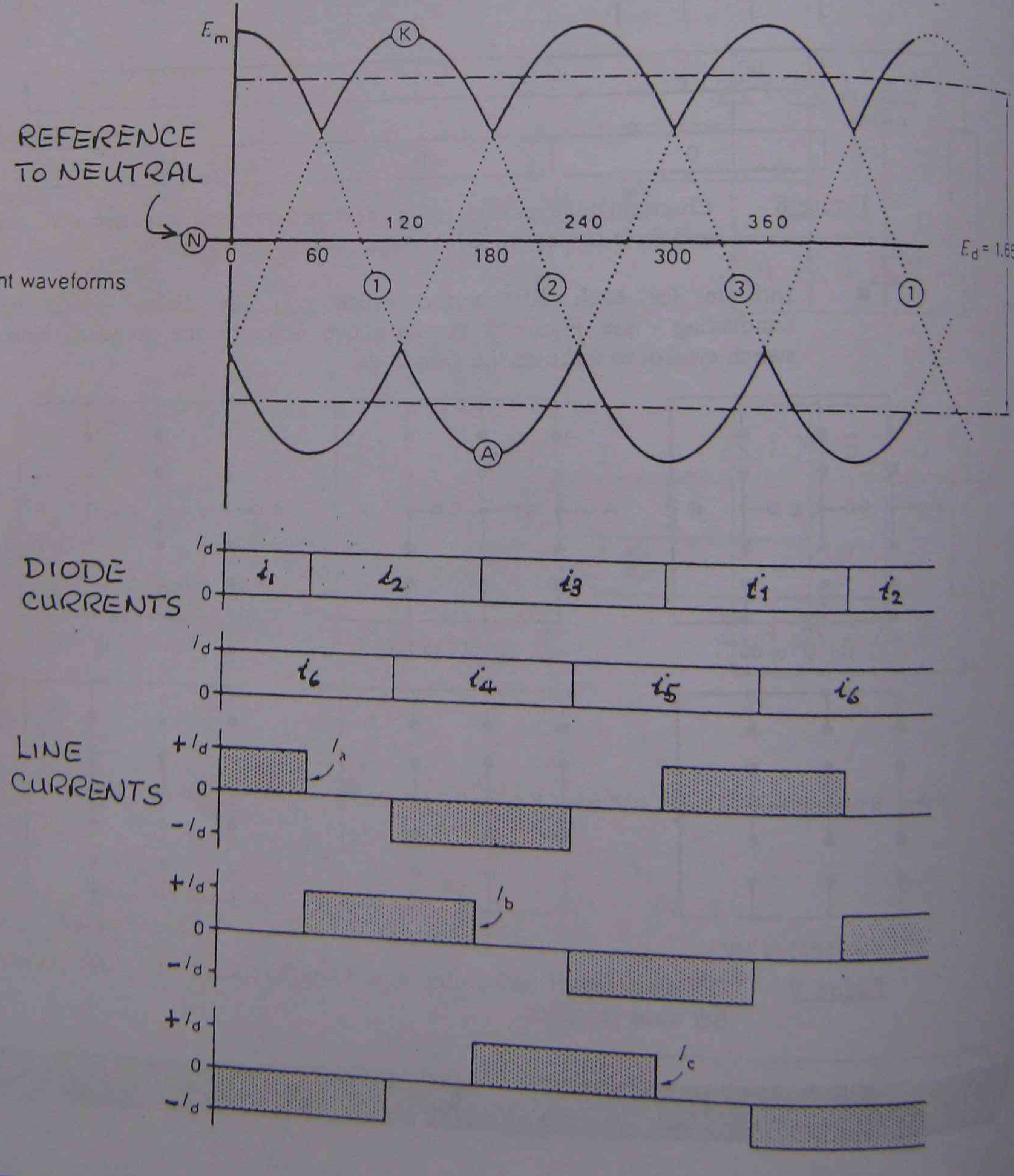


Figure 21-20
Voltage and current waveforms
in Fig. 21-19.

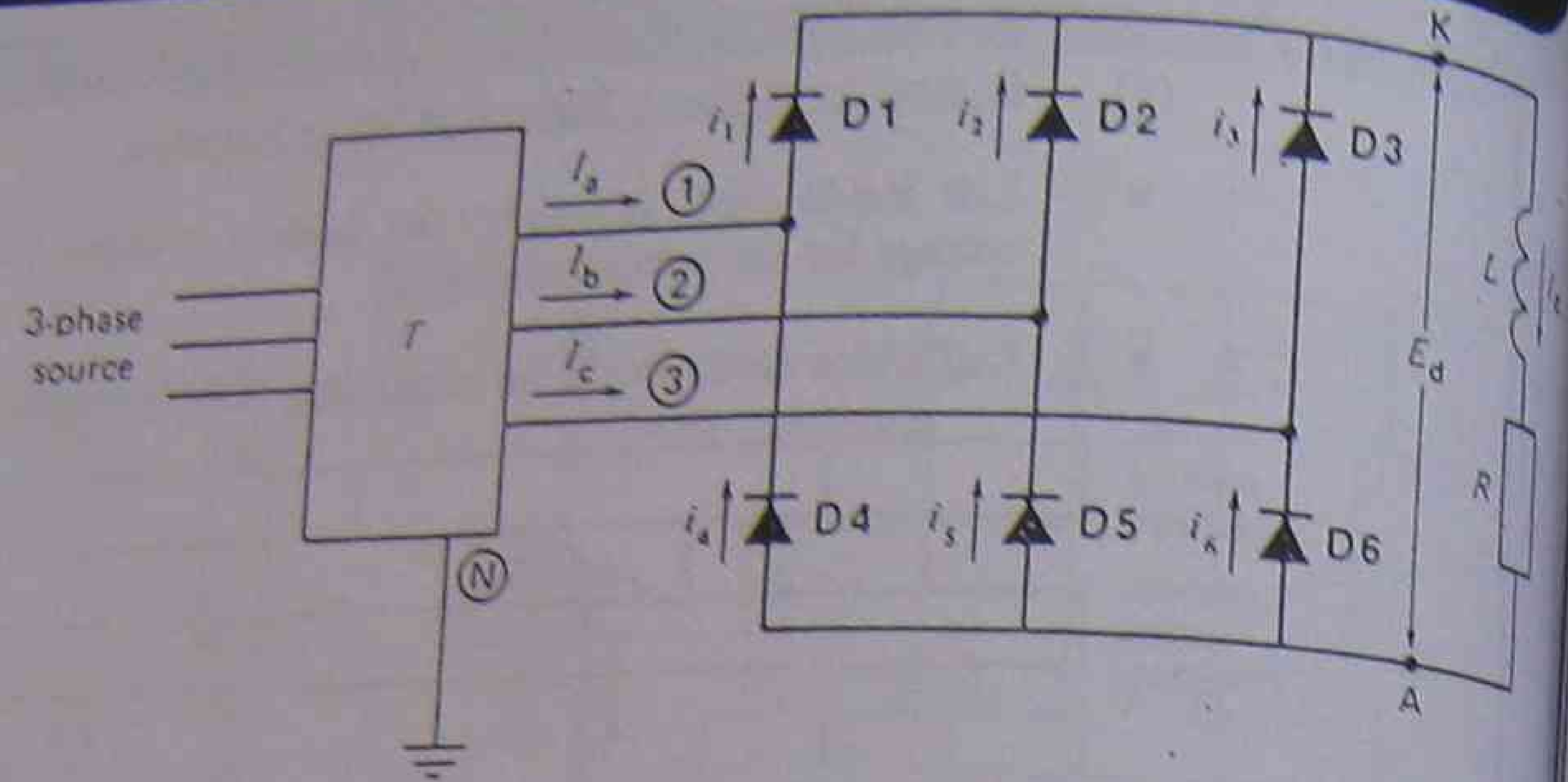


Figure 21-19
Three-phase, 6-pulse rectifier with inductive filter

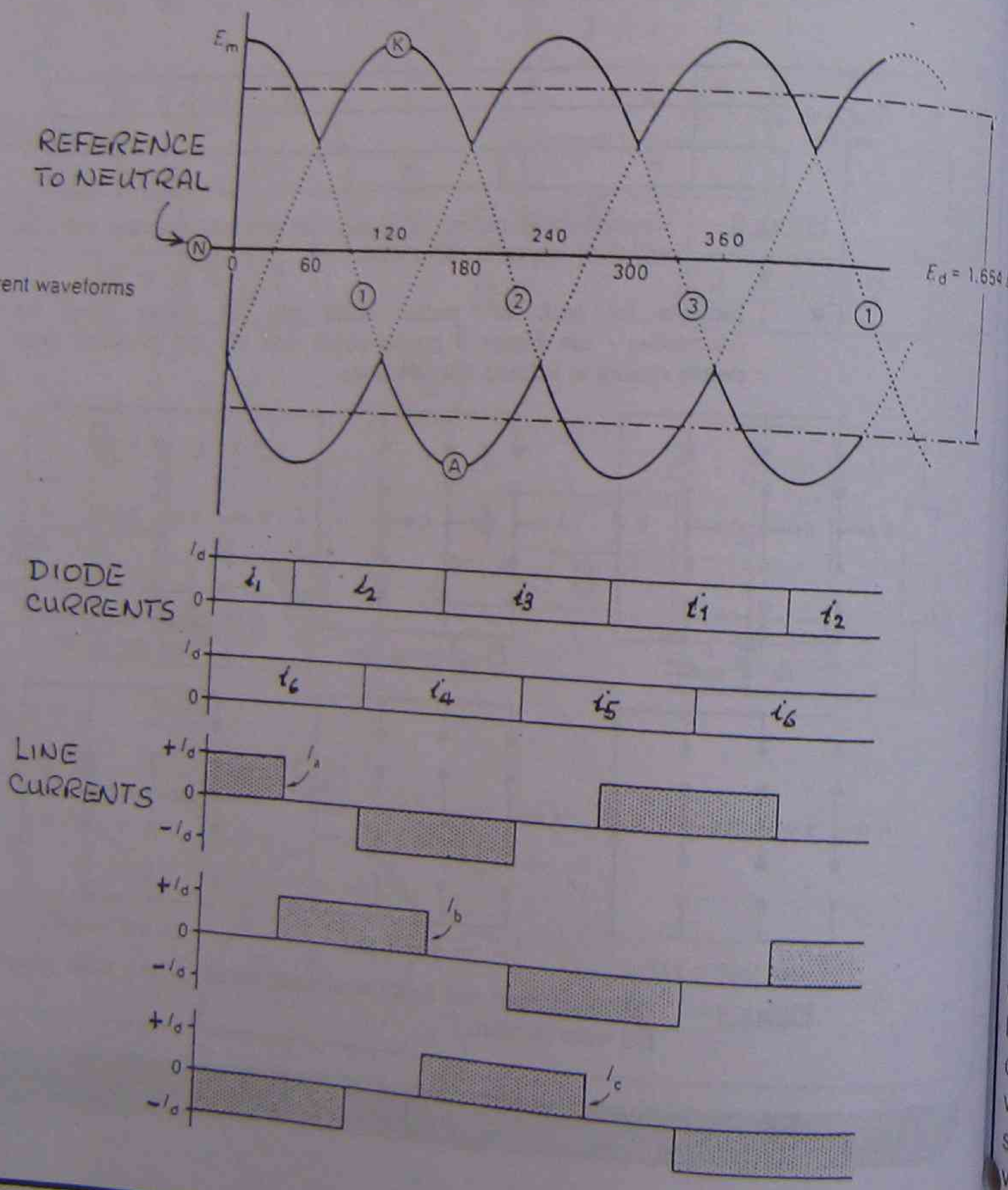


Figure 21-20
Voltage and current waveforms in Fig. 21-19.

21.26 Basic principle of operation

We can gain a basic understanding of how the converter works in the rectifier mode by referring to Fig. 21-48. In this figure, the SCRs are assumed to be enclosed in a box, where they successively switch the output terminals K, A to the ac supply lines 1, 2, 3. The load is represented by a resistor in series with an inductor L. The inductor is assumed to have a very large inductance, so that the load current I_d remains constant. In Fig. 21-48a, the two thyristors Q1, Q5 located between terminals K-1 and A-2 are conducting. A moment later, the thyristors Q2, Q4 between K-2 and A-1 conduct (Fig. 21-48b). The other thyristors are similarly switched, in sequence. When these steps have been completed, the entire switching cycle repeats. The reader will note that the dc current I_d flows in the ac lines. However, Fig. 21-48 shows that the current in each line reverses periodically, and so it is a true ac current of amplitude I_d. It is also evident that the current in a particular line is zero for brief intervals. Thus, there is momentarily no current in line 3 of Fig. 21-48.

The switching sequence we have just described is similar to that of the diode bridge rectifier of Fig. 21-22. There is, however, an important difference. The thyristors can be made to conduct at precise moments on the ac voltage cycle. Thus, conduction can be initiated when the instantaneous voltage between the ac lines is either high or low. If the voltage is low, the dc output voltage will also be low. Conversely, if the thyristors conduct when the ac line voltage is momentarily near its peak, the dc output voltage will be high. In effect, the output voltage E_{KA} is composed of short 60-degree segments of the ac line voltage. The average value of E_{KA} is the dc output voltage E_d.

In examining Fig. 21-48, it can be seen that the line current always flows out of a line that is momentarily positive. This must be so because the line delivers active power to the load. For example, in Fig. 21-48a, e_{12} is positive when I_d flows in the direction shown.

✳ Now that we know how the thyristor converter behaves as a rectifier, the question arises; how can it be made to operate as an inverter? Three basic conditions have to be met.

• First, we must have a source of dc current I_d. Such a current source can be provided if a voltage source E_0 is connected in series with a large inductance (Fig. 21-49a).

• Second, the converter must be connected to a 3-phase line that can maintain an undistorted sinusoidal voltage, even when the line current is nonsinusoidal. The voltage may be taken from a power utility, or generated by a local alternator.

• Third, the thyristors must be switched so that current I_d flows into an ac line that is momentarily positive. The gate firing must therefore be precisely synchronized with the line frequency.

The inverter operation can best be understood by referring to Fig. 21-49. The SCRs enclosed in the box are arranged the same way as in Fig. 21-48. In other words, the converters in the two figures are absolutely identical. Looking first at the dc side, the dc current I_d must flow in the same direction as before because SCRs cannot conduct in reverse. On the other hand, because we want the dc source E_0 to deliver power, I_d must flow out of the positive terminal, as shown. In other words, the positive side of E_0 must be connected to terminal A.

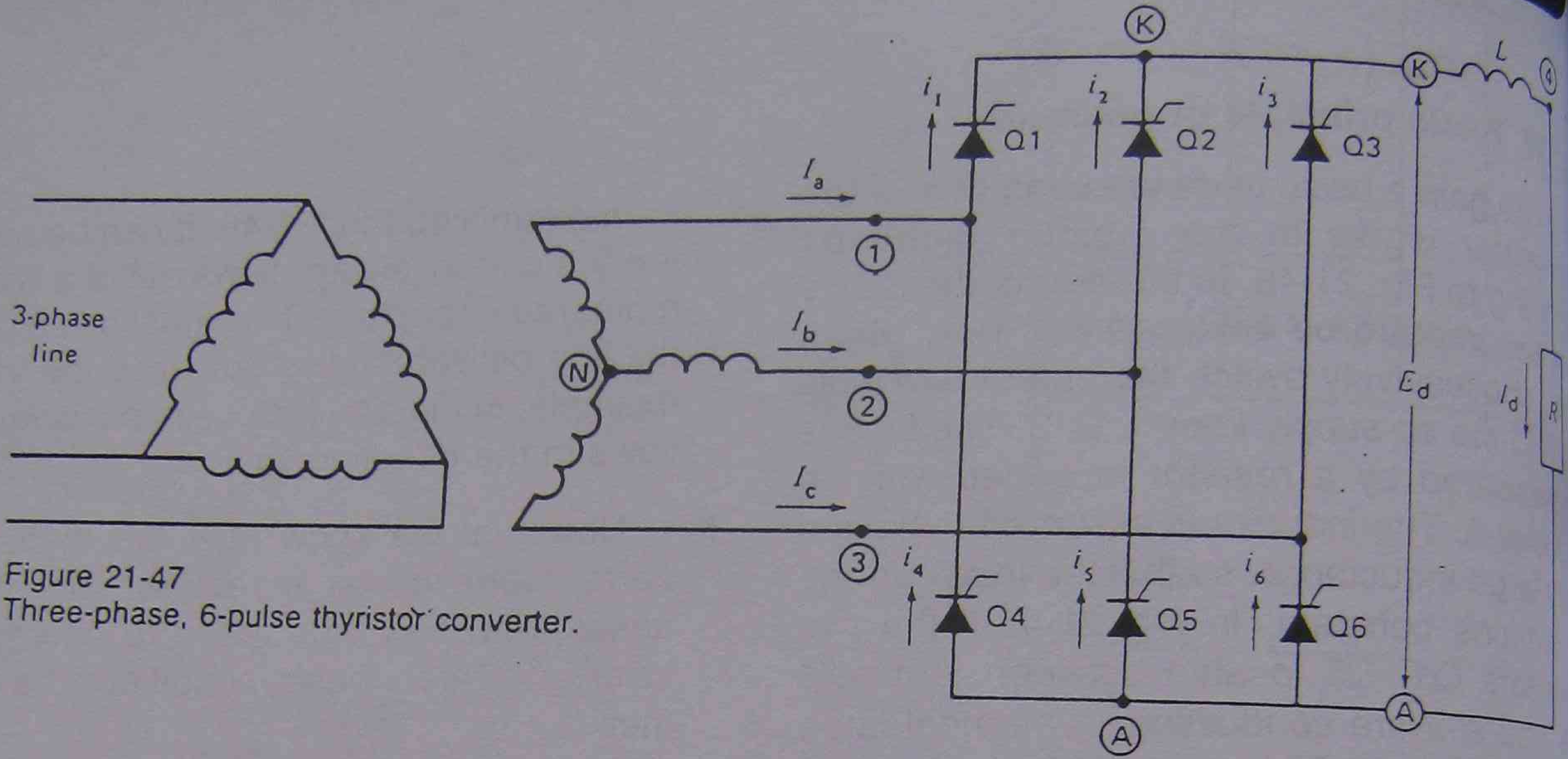


Figure 21-47
Three-phase, 6-pulse thyristor converter.

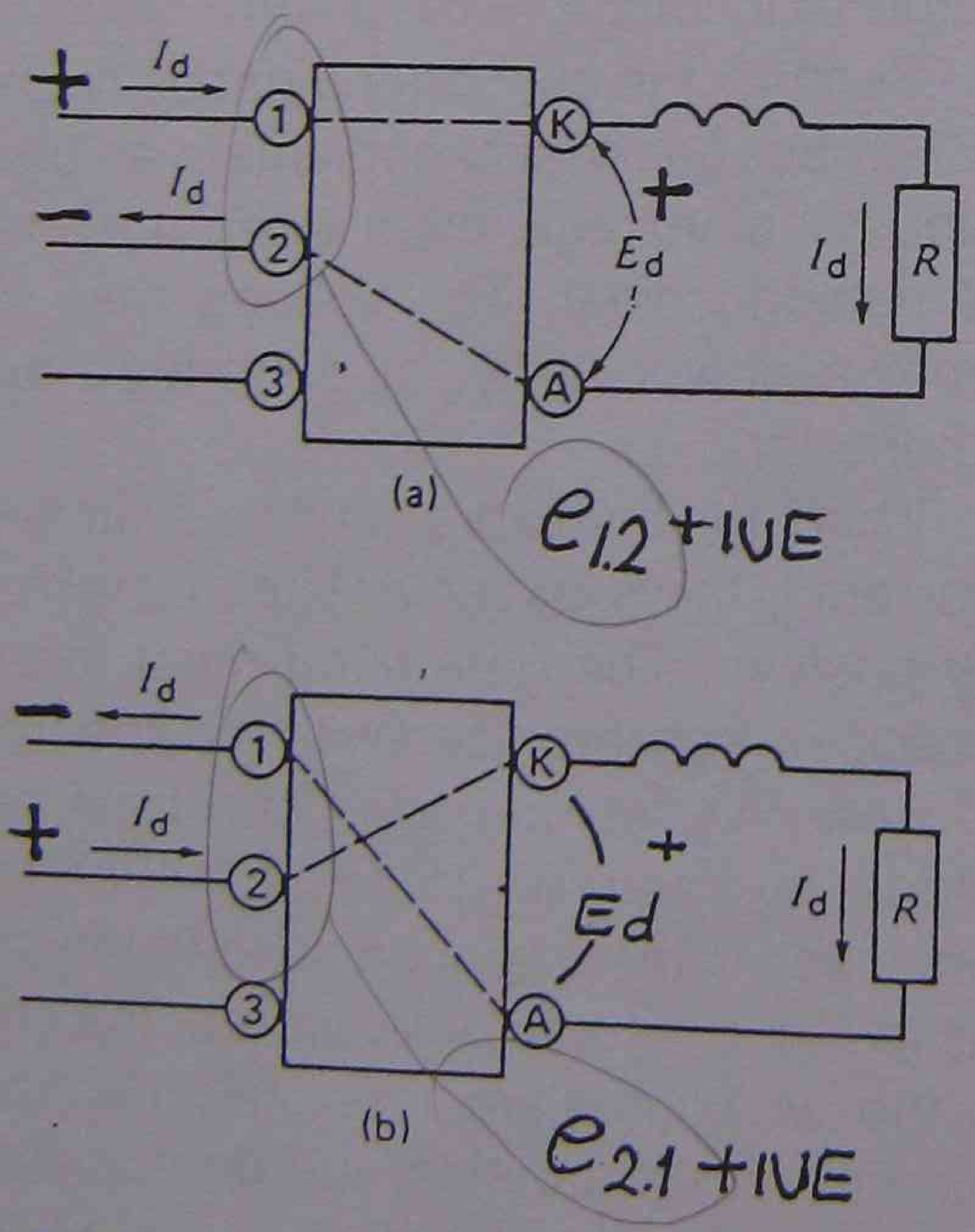


Figure 21-48
Rectifier mode (see Fig. 21-47)
a. Q1 and Q5 conducting.
b. Q2 and Q4 conducting.

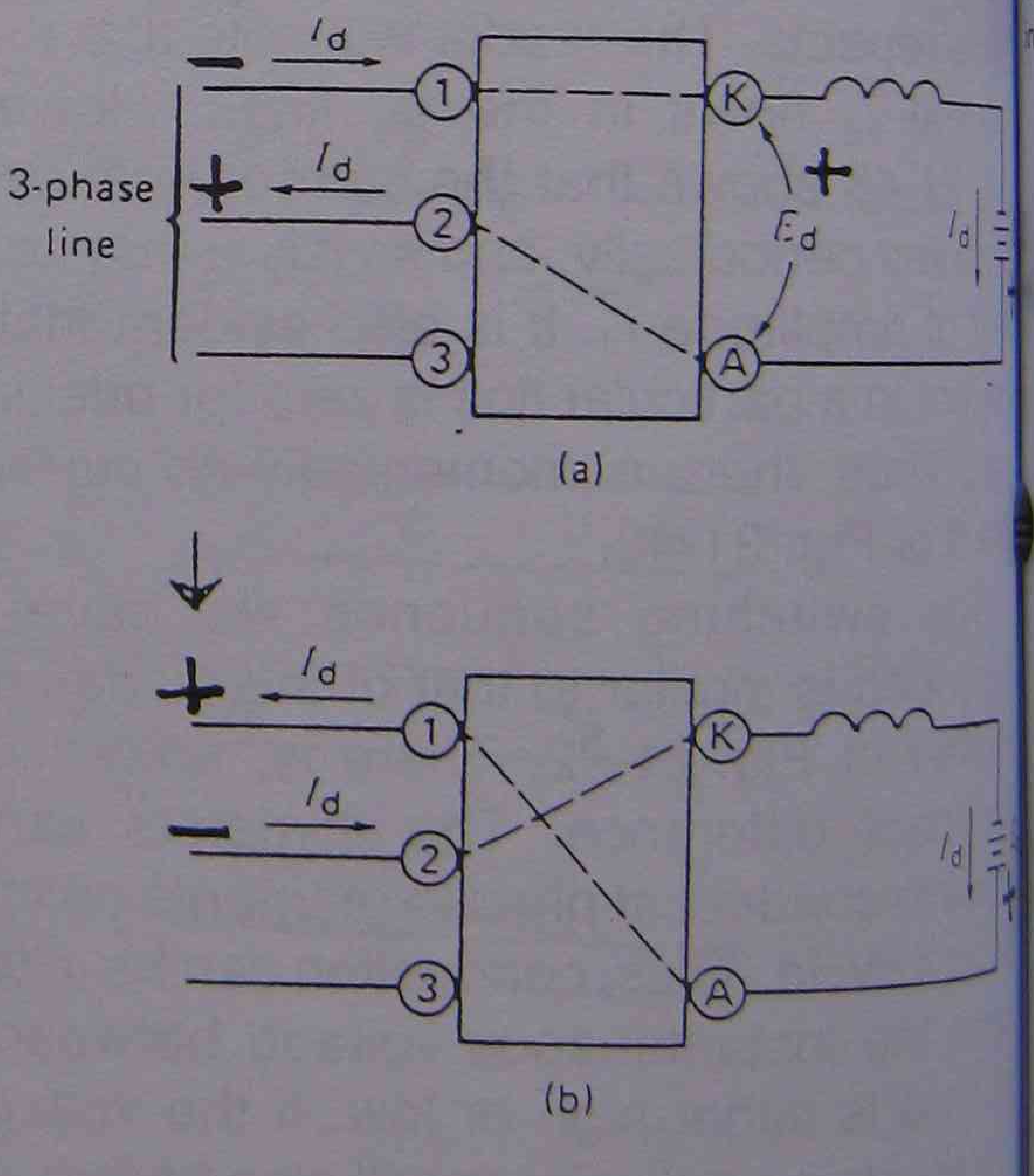


Figure 21-49
Inverter mode (see Fig. 21-47)
a. Q1 and Q5 conducting.
b. Q2 and Q4 conducting.

Because we can initiate conduction whenever we please, the thyristors enable us to vary the dc output voltage when the converter operates in the rectifier mode. The converter can also function as an inverter, provided that a dc source is used in place of the

10

A.C. Motor Control.

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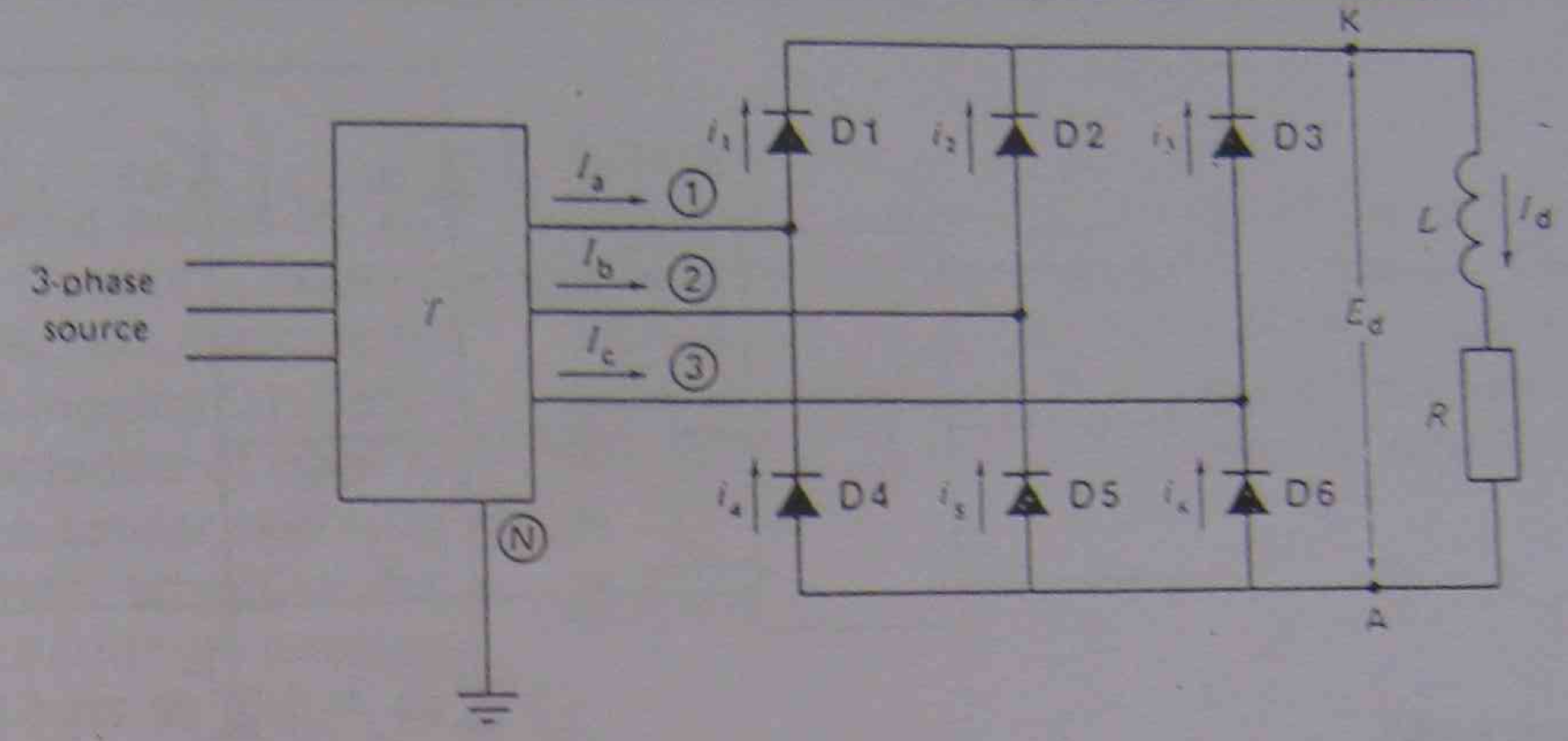

Figure 21-19
Three-phase, 6-pulse rectifier
with inductive filter.

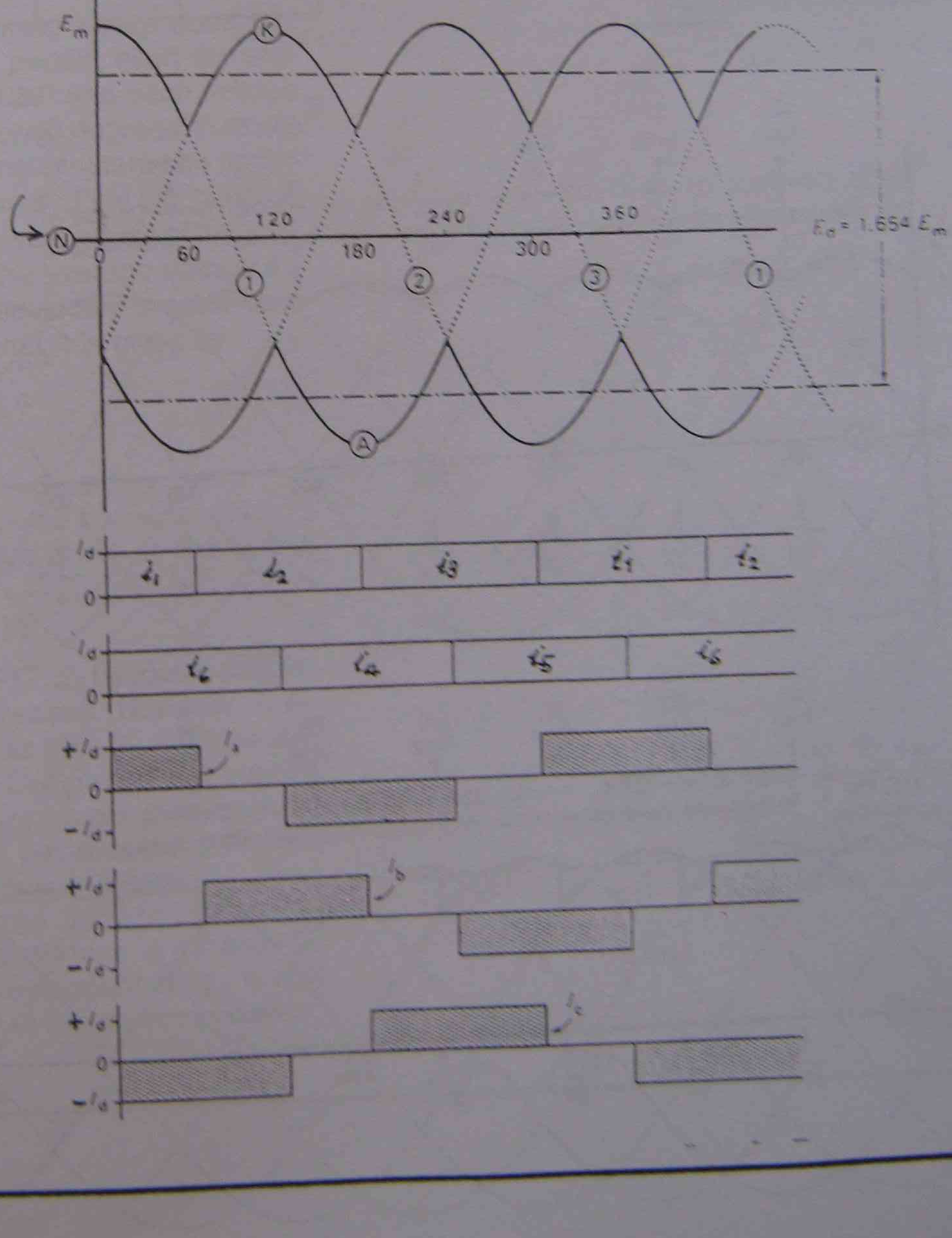

Figure 21-20
Voltage and current waveforms
in Fig. 21-19.

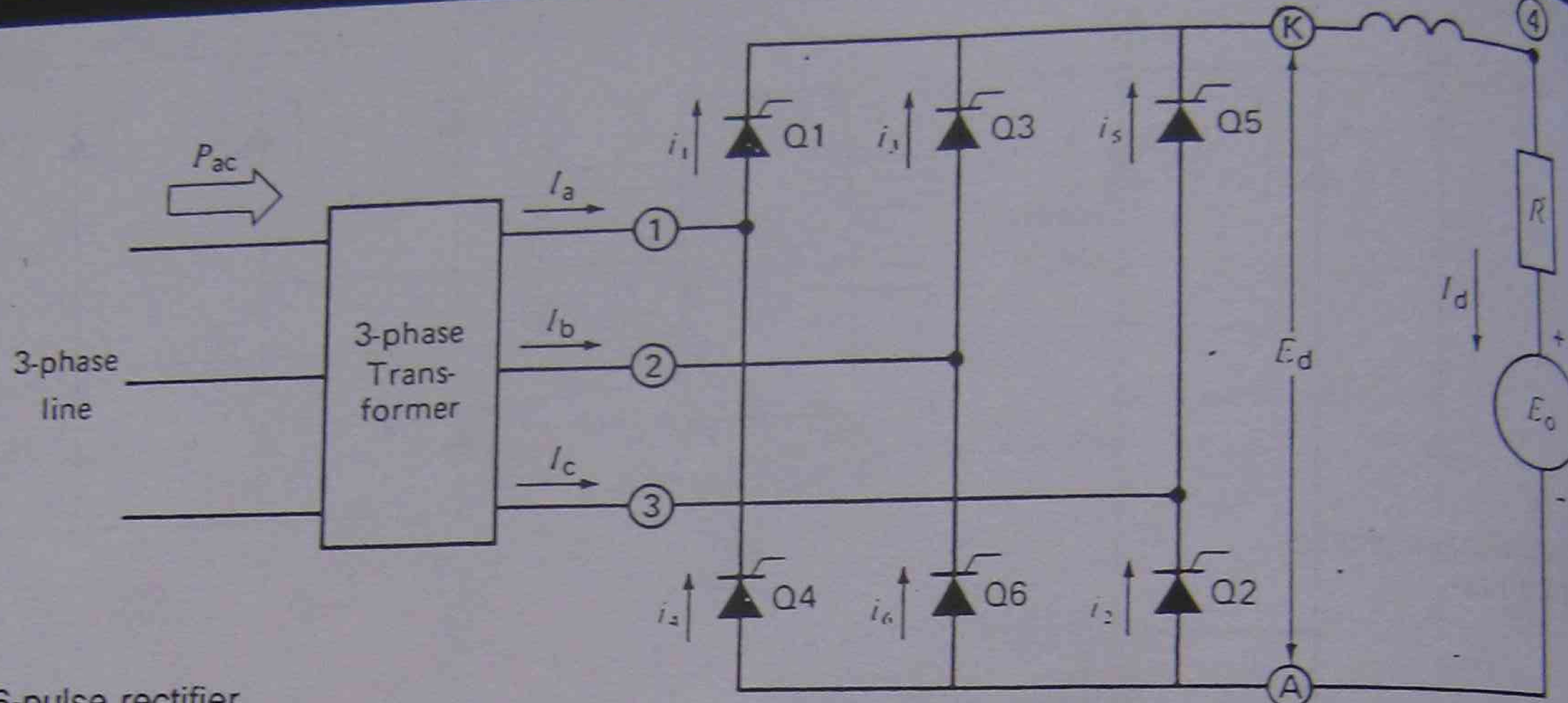

Figure 21-50
Three-phase, 6-pulse rectifier.

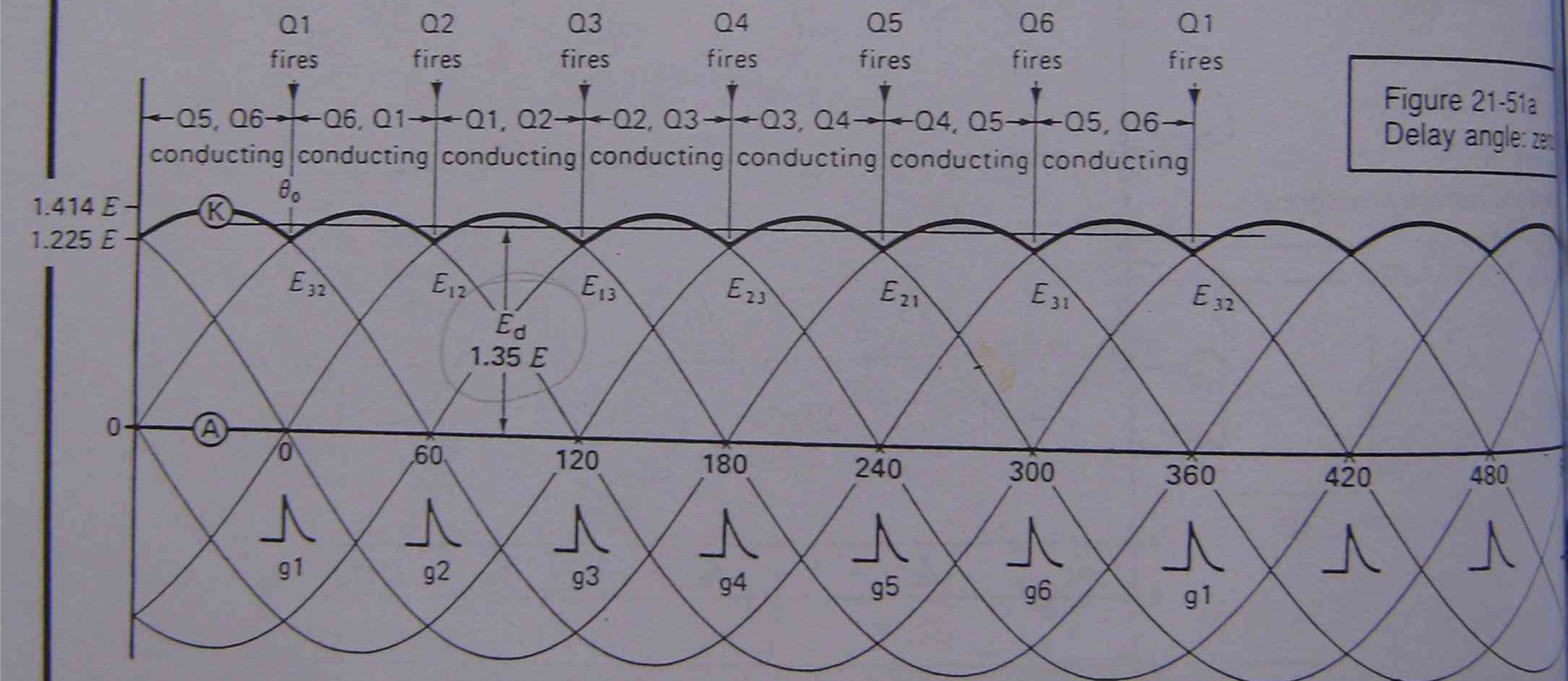

Figure 21-51a
Delay angle: zero

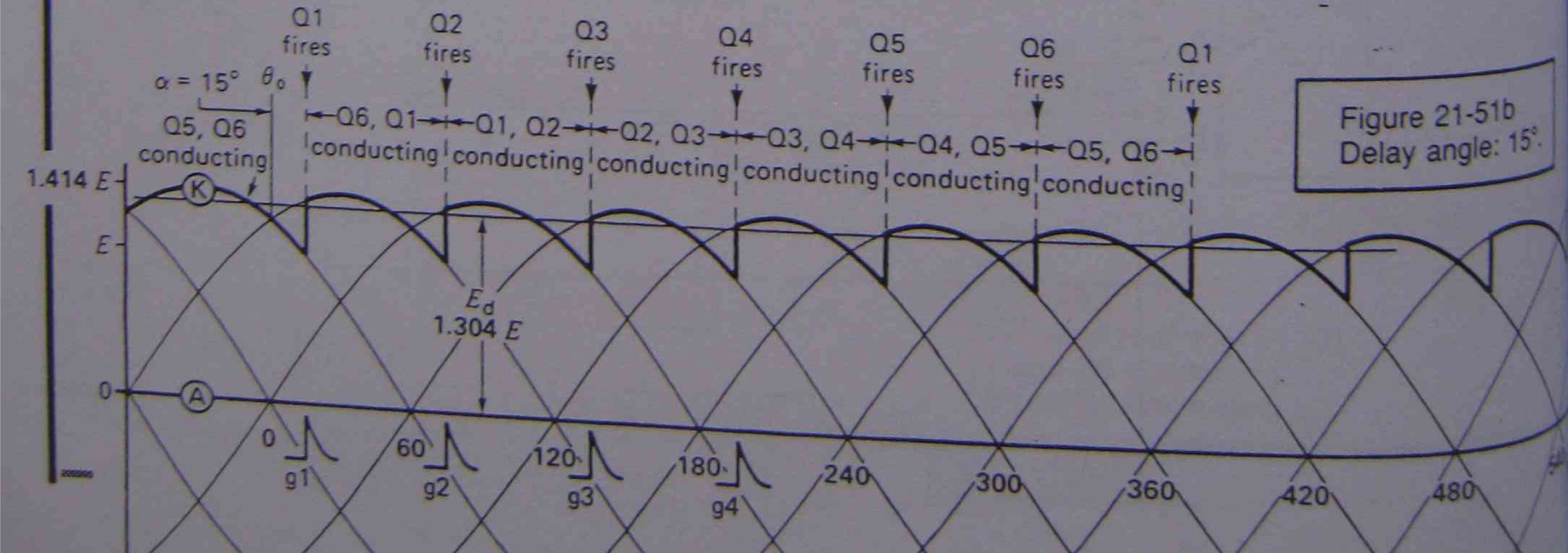

Figure 21-51b
Delay angle: 15°.

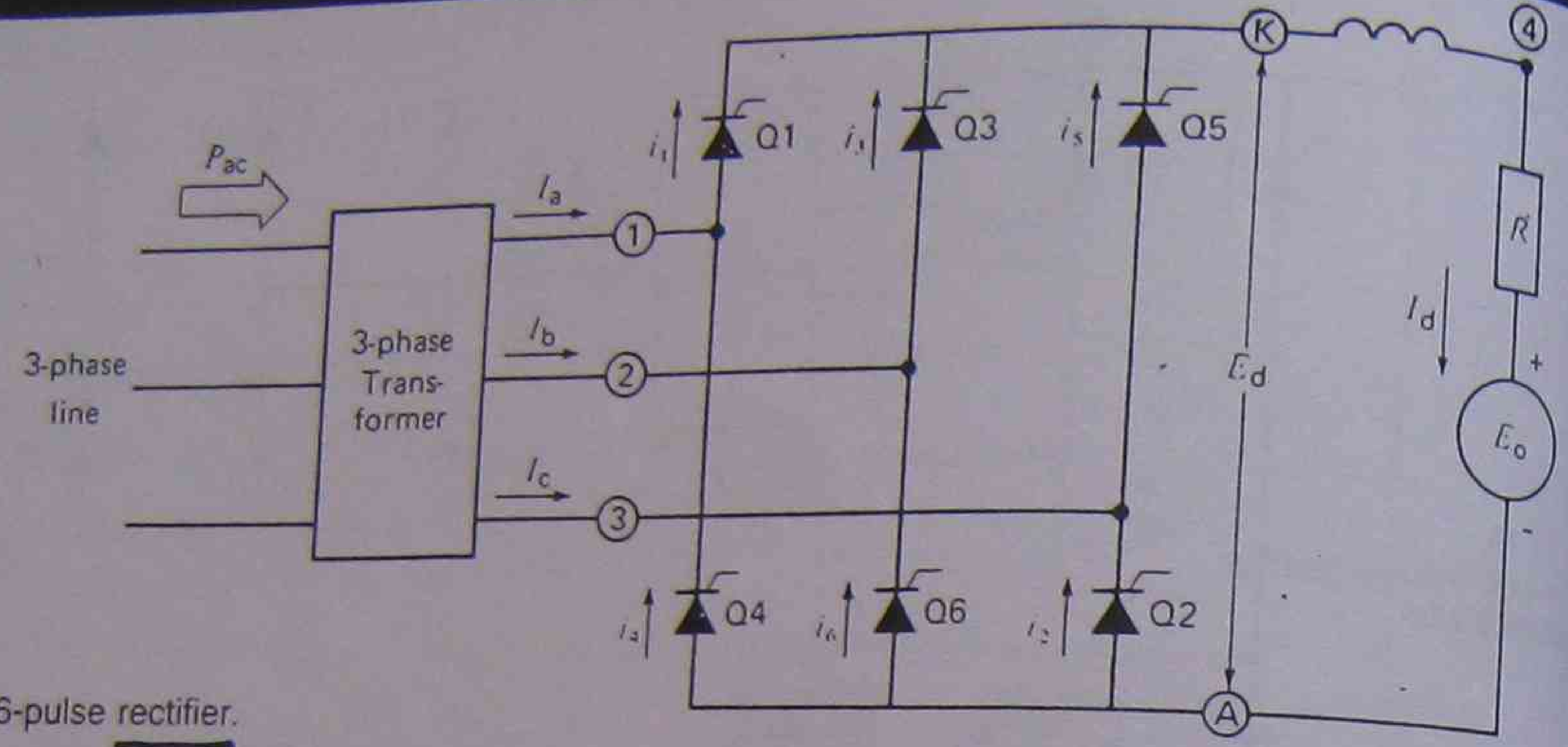

Figure 21-50
Three-phase, 6-pulse rectifier.

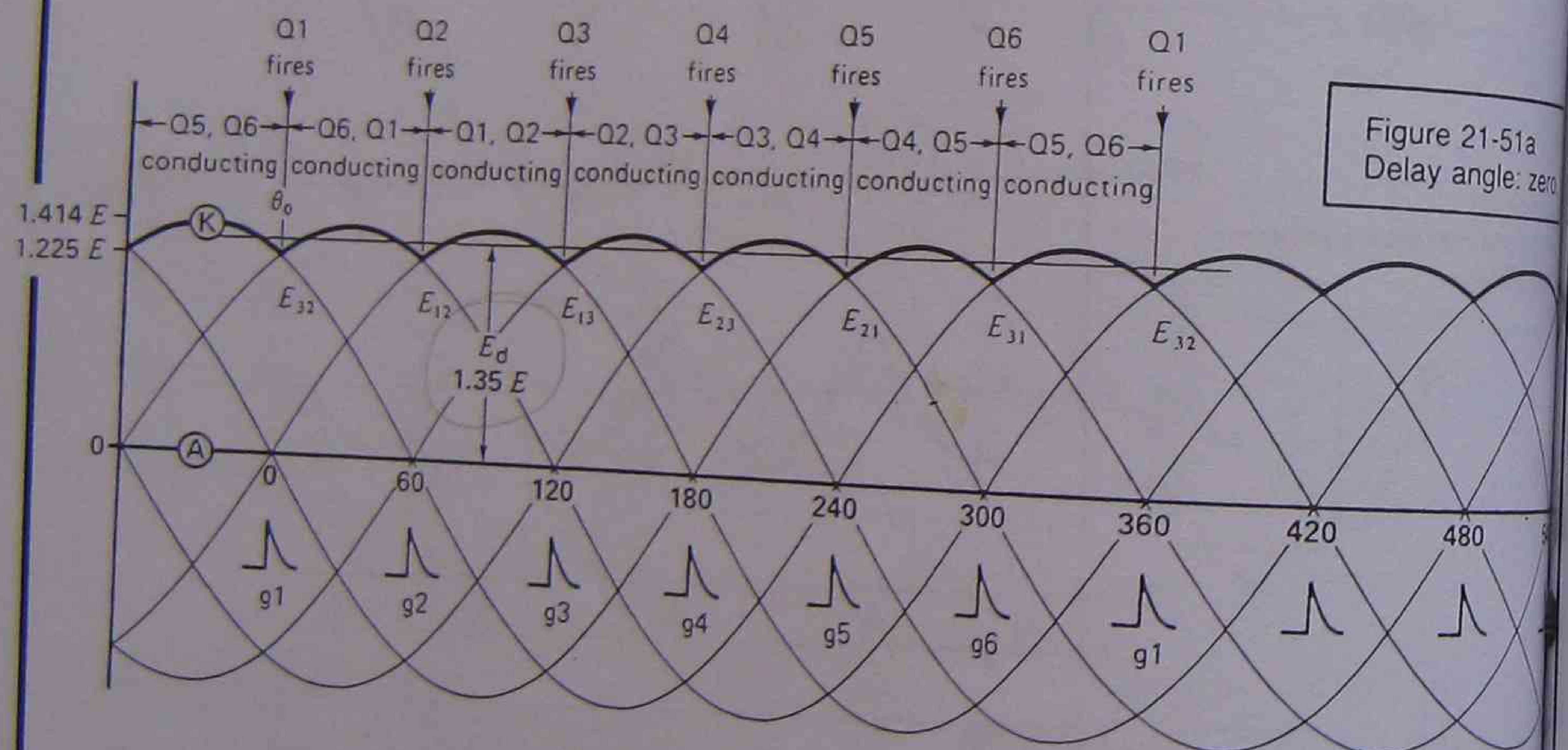

Figure 21-51a
Delay angle: zero

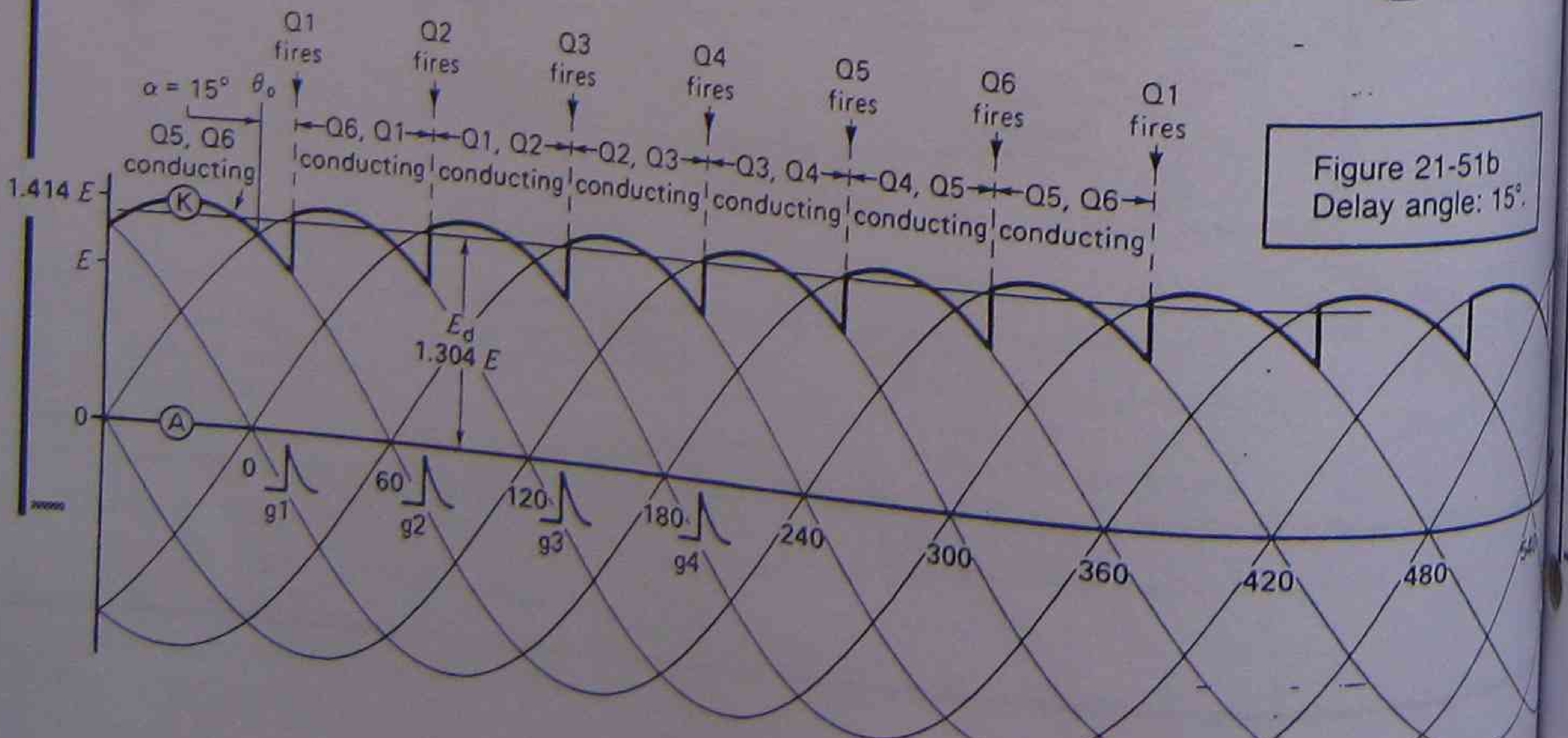

Figure 21-51b
Delay angle: 15°.

21.28 Delayed triggering – rectifier mode

Let us now delay all triggering pulses by an angle α of 15° (Fig. 21-51b). Current I_d, instead of switching over to Q1 at θ_0, will continue to flow in Q5 until gate pulse g1 triggers Q1. Commutation occurs, and the potential of point K jumps from line 3 to line 1. A similar switching action takes place (but at later times) for the other thyristors. The resulting choppy waveshape between terminals K and A is shown in Fig. 21-51b.

Note that the triggering delay does not shorten the conduction period; each thyristor still conducts for a full 120° and each voltage segment has a duration of 60 degrees. Furthermore, the current remains constant and ripple-free, owing to the presence of the big inductor. The level of point K follows the tops of the individual sine waves, but the average voltage E_d, between K and A, is obviously smaller than before. We can prove that it is given by:

$$E_d = 1.35E\cos\alpha \qquad (21\text{-}17)$$

where

E_d = dc voltage produced by the 3-phase, 6-pulse converter [V]
E = effective value of the ac line-to-line voltage [V] — Vrms
α = firing angle [°]

According to Eq. 21-17, E_d becomes smaller and smaller as α increases. However, if E_d becomes equal to or less than E_0, the load current ceases to flow. Ordinarily, the current would reverse when E_d is smaller than E_0. However, this is impossible, because the SCRs can only conduct in the forward direction.

Figures 21-51c and 21-51d show the waveform between K and A for α = 45° and 75°, respectively. The ac component in E_{KA} is now very large, compared to the dc component.

Example 21-9.
The 3-phase converter of Fig. 21-50 is connected to a 3-phase 480 V, 60 Hz source. The load consists of a 500 V dc source having an internal resistance of 2 Ω. Calculate the power supplied to the load for triggering delays of a. 15°, b. 75°.

Solution:

③ Active Power $P_1 = S_1 \cos\phi$ kW
$P_1 = 1.35 V_{RMS} I_D \cos\phi$ kW

② Apparent Power $S_1 = \sqrt{3} V_{rms} I_1$ KVA
$= \sqrt{3} V_{rms}\, 0.78 I_D$ KVA
$= 1.35 V_{rms} I_D$

④ Reactive Power $Q = S_1 \sin\phi$ kVAr

① $I_{1\,rms} = \sqrt{3}\frac{\sqrt{2}}{\pi} I_D = 0.78 I_D$ amps
I_D = DC current.

Delay angle	converter behaviour
α = 0°	Resistive load
0° < α < 90°	Resistive/Inductive load
α = 90°	Pure Inductive load
α > 90°	Source from Ind load

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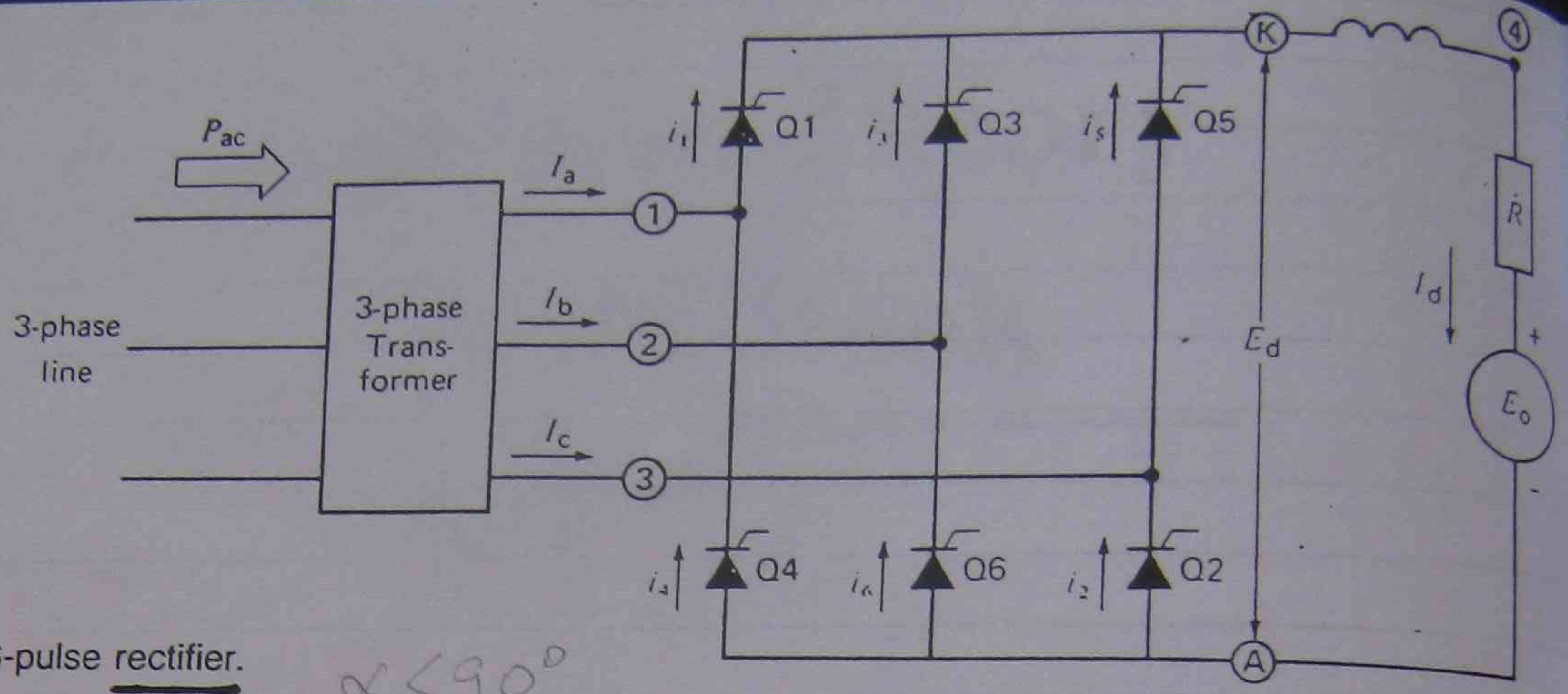


Figure 21-50
Three-phase, 6-pulse rectifier.

$\alpha < 90°$

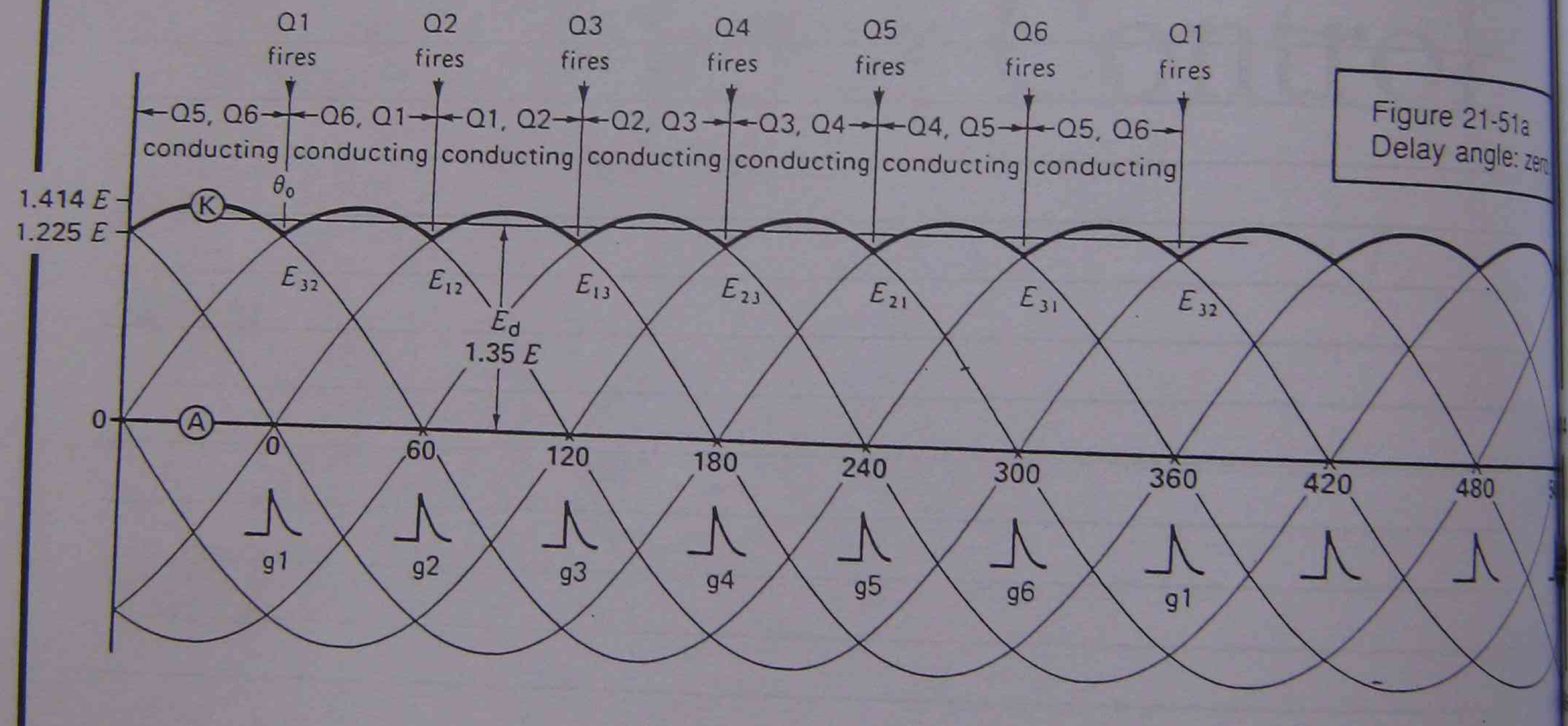


Figure 21-51a
Delay angle: zero

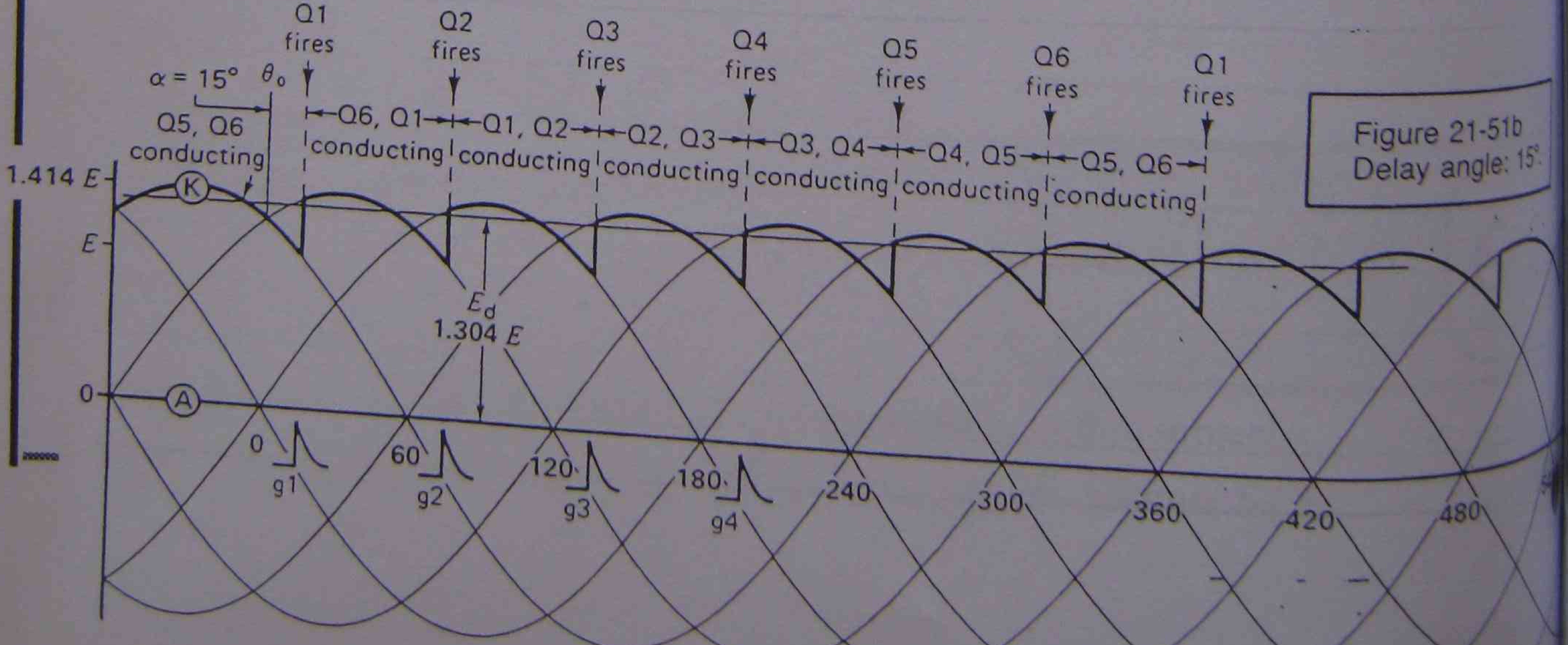


Figure 21-51b
Delay angle: 15°

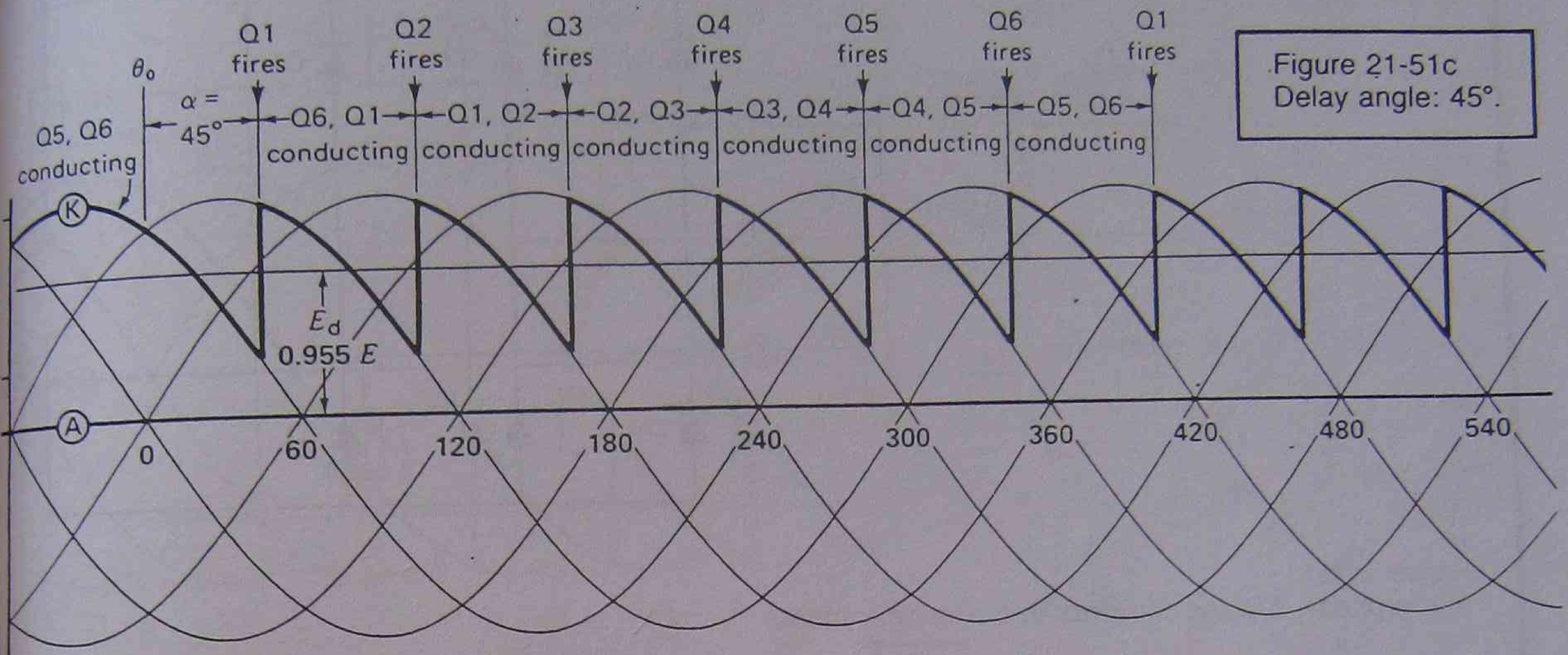


Figure 21-51c
Delay angle: 45°.

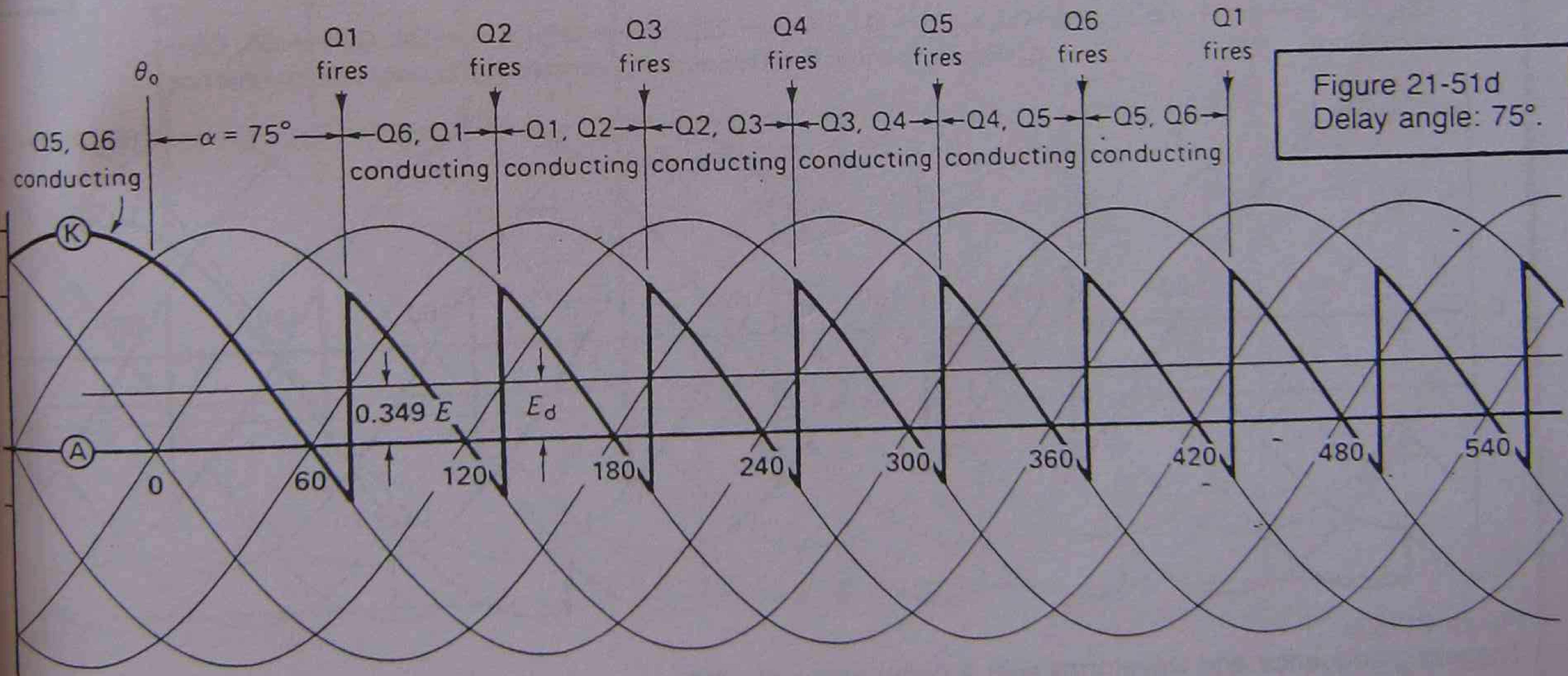
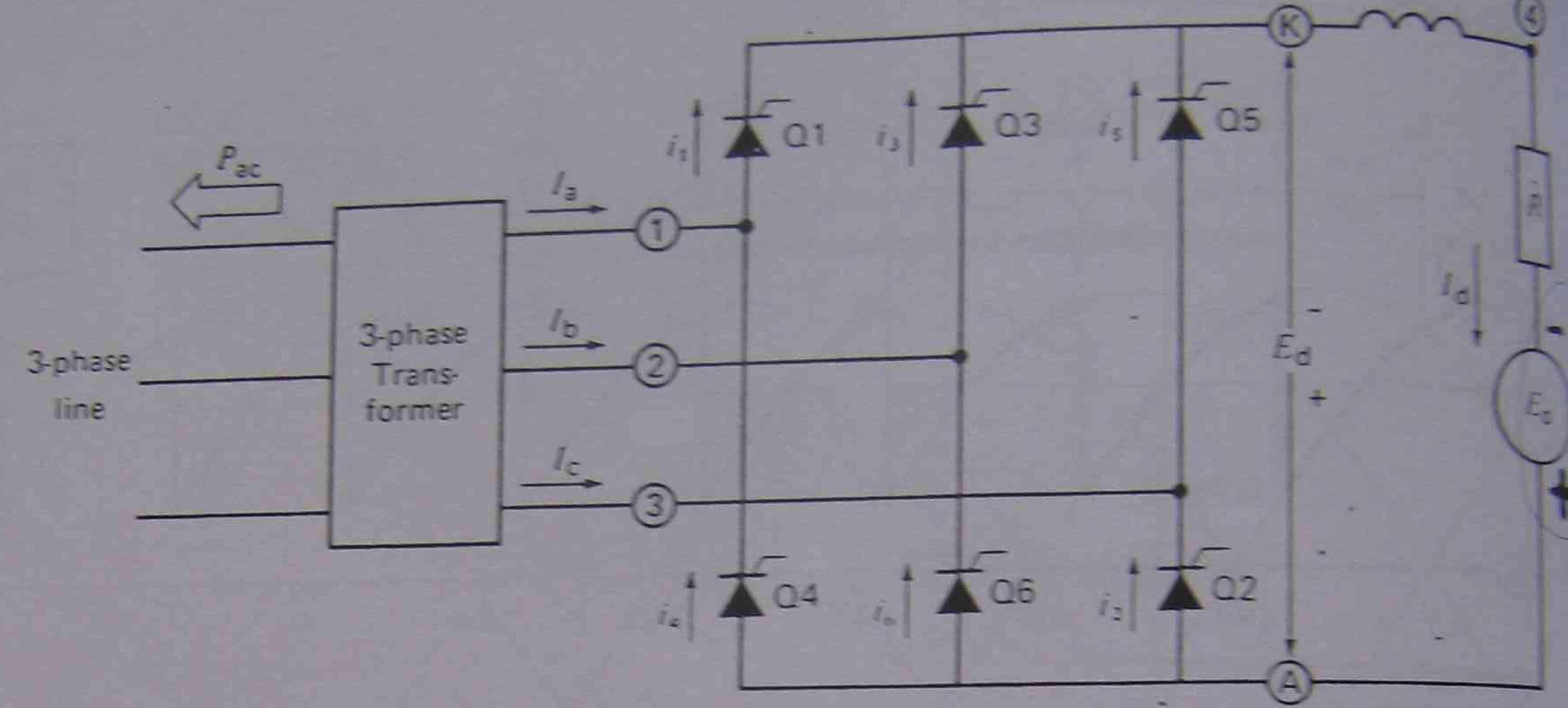


Figure 21-51d
Delay angle: 75°.



note E_0 polarity
i.e. BEMF

Figure 21-52
Three-phase, 6-pulse converter in the inverter mode.

α > 90°

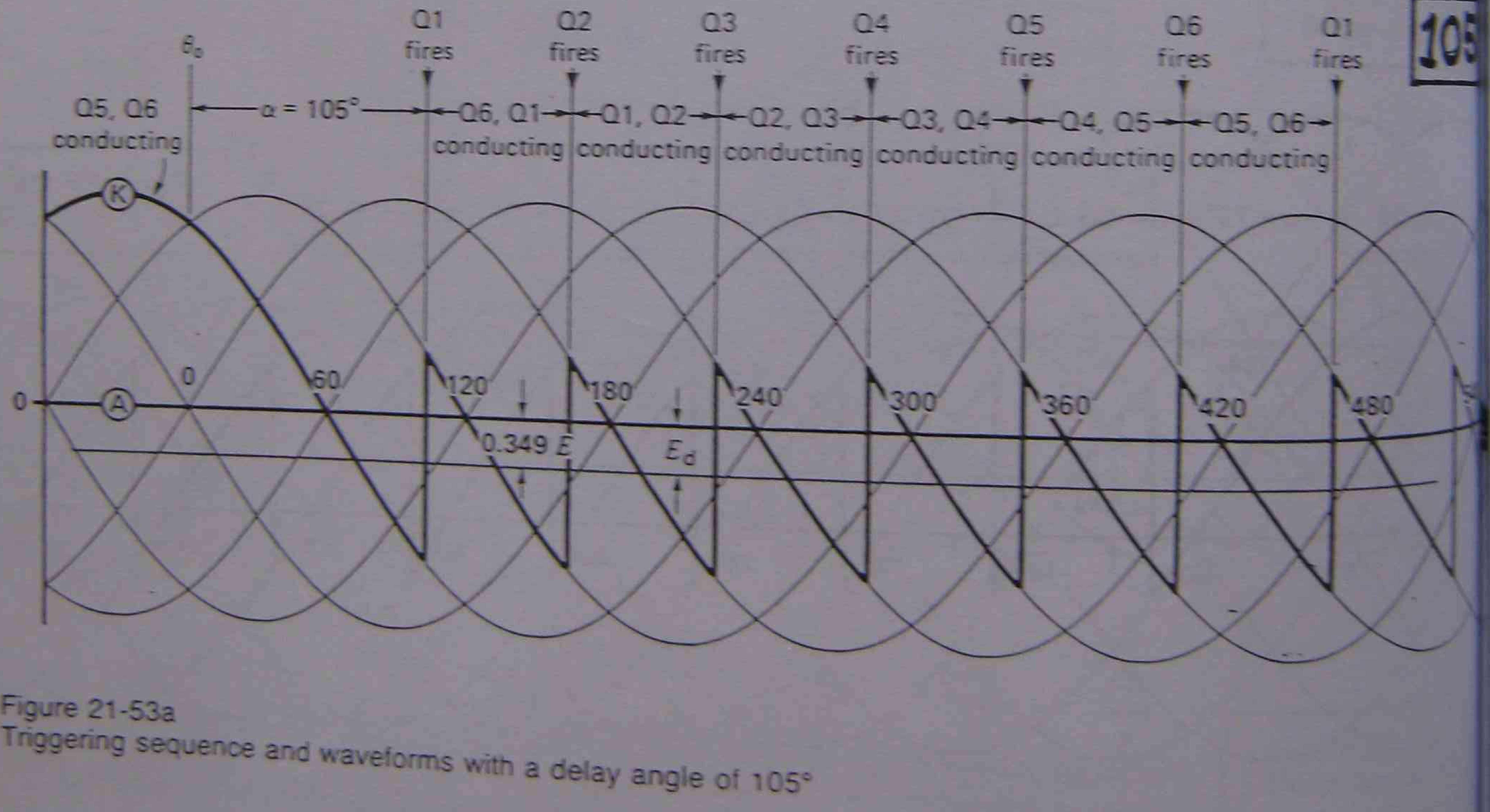


Figure 21-53a
Triggering sequence and waveforms with a delay angle of 105°

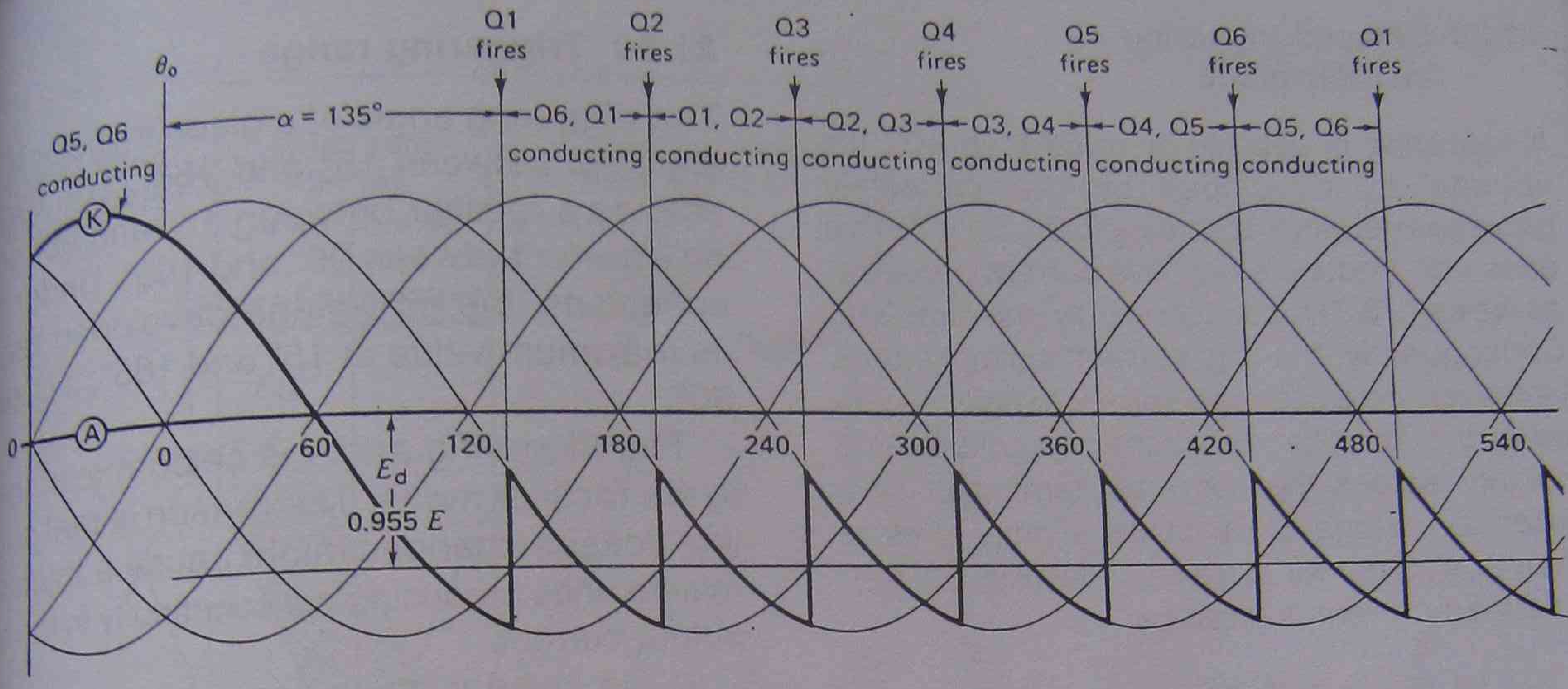


Figure 21-53b
Triggering sequence and waveforms with a delay angle of 135°

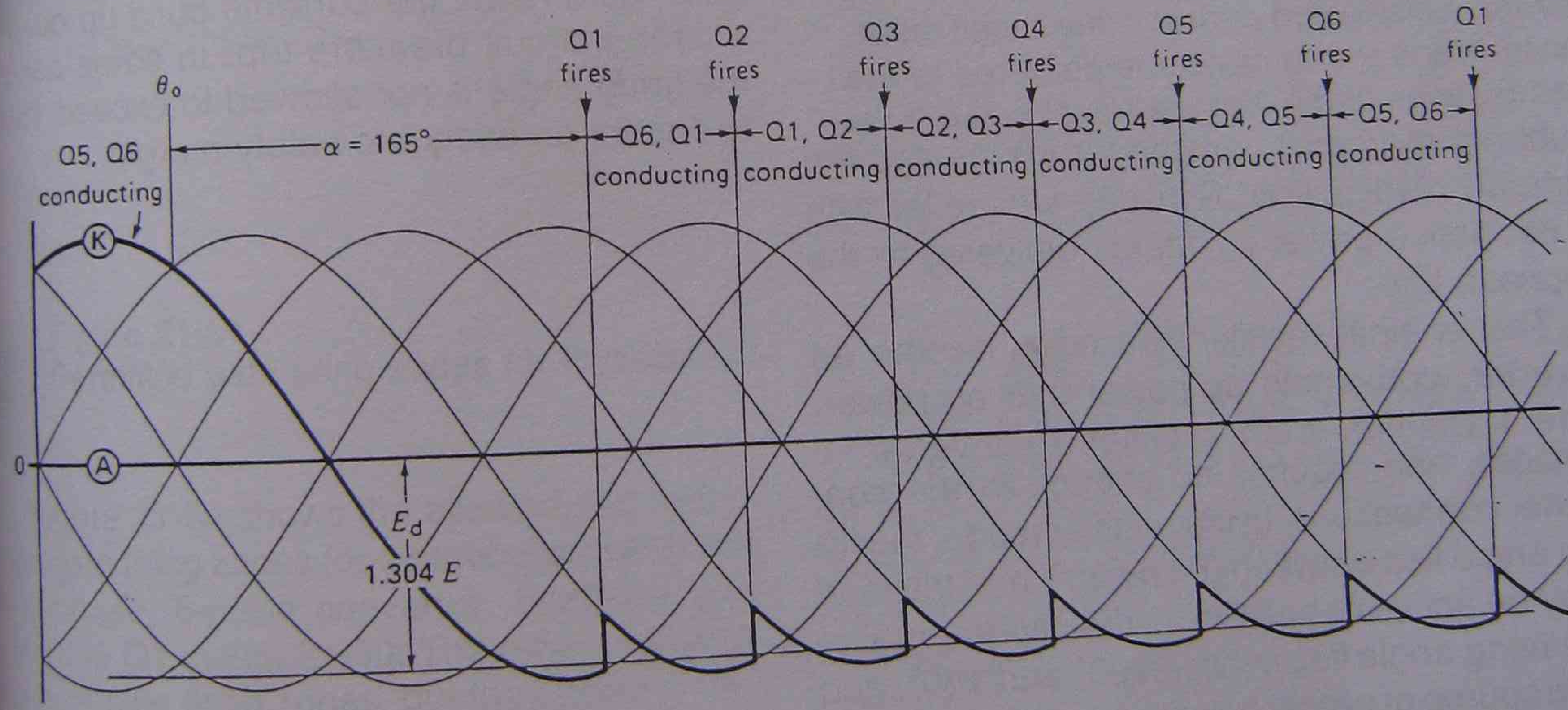


Figure 21-53c
Triggering sequence and waveforms with a delay angle of 165°.

21.29 Delayed triggering – inverter mode

If triggering is delayed by more than 90°, the voltage E_d developed by the converter becomes negative according to Eq. 21-17. This does not produce a negative current because, as we said, SCRs conduct in only one direction. Consequently, the load current is simply zero. However, we can *force* a current to flow by connecting a dc voltage of proper magnitude and polarity across the converter terminals. This external voltage E_0 must be slightly greater than E_d in order for current to flow (Fig. 21-52). The load current is given by:

$$I = (E_0 - E_d)/R$$

Because current flows out of the positive terminal of E_0, the "load" is actually a source, delivering a power output $P = E_0 I_d$. Part of this power is dissipated as heat in the circuit resistance R and the remainder is delivered to the secondaries of the 3-phase transformer. If we subtract the small transformer losses and the virtually negligible SCR losses, we are left with a net active power P_{ac} that is delivered to the 3-phase line.

The original rectifier has now become an *inverter*, converting dc power into ac power. The transition from rectifier to inverter is smooth, and requires no change in the converter connections. In the rectifier mode, the firing angle lies between 0° and 90°, and the load may be active or passive. In the inverter mode, the firing angle lies between 90° and 180°, and a dc source of proper polarity *must* be provided.

Figure 21-53 shows the waveshapes at firing angles of 105°, 135° and 165°. The dc voltage E_d generated by the inverter is still given by Eq. 21-17. It reaches a maximum value of $E_d = -1.35E$ at a firing angle of 180°.

21.30 Triggering range

The triggering angle of a given thyristor is u ally kept between 15° and 165°. The thyris acts as a rectifier between 15° and 90° and an inverter between 90° and 165°. Under th conditions, the dc voltage developed rea its maximum value at 15° and 165°; it is zer 90°.

The triggering angle is seldom less than in the rectifier mode. The reason is that sudd line voltage changes might cause a thyristor misfire, thus producing a discontinuity in the dc output current.

In the inverter mode, we never permit the firing angle to exceed 165°. If we go beyond this point, the inverter begins to lose its ability to switch reliably from one thyristor to the next. As a result the currents build up quickly until the circuit breakers trip. In some cases the firing angle is not allowed to exceed 150° to ensure an adequate safety margin.

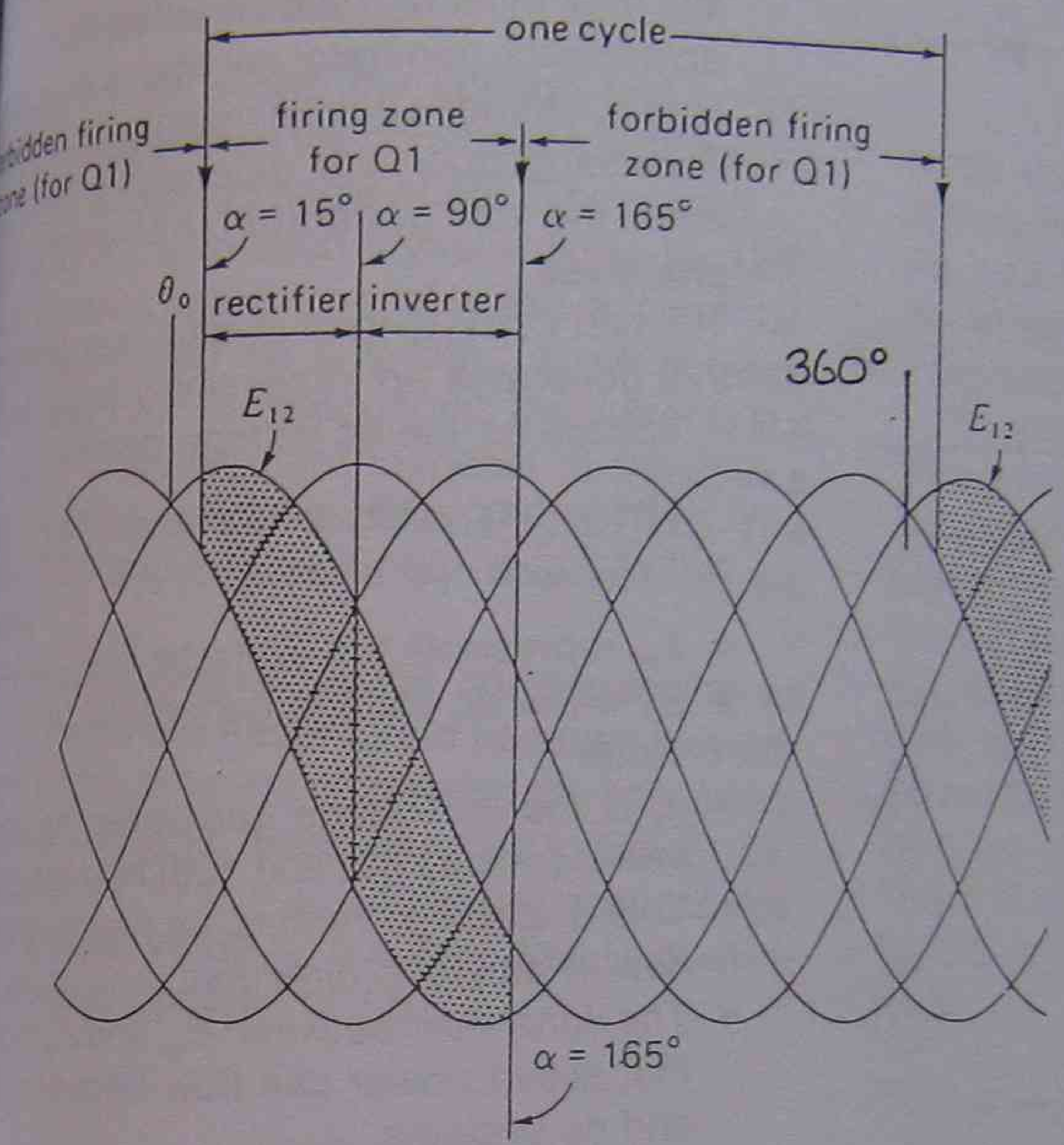
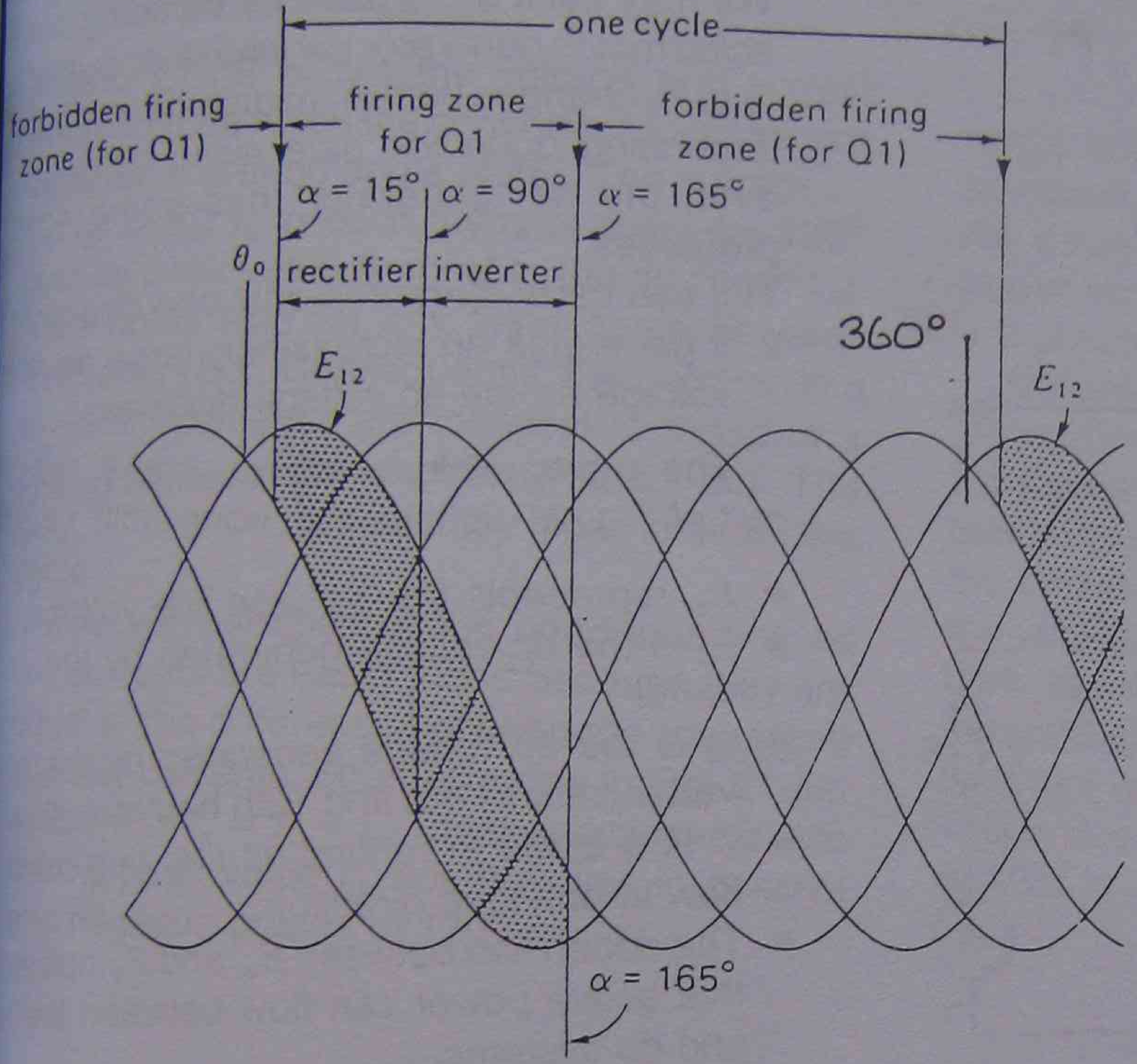


Figure 21-54
Permitted gate firing zones for thyristor Q1.

Figure 21-54 shows the allowed and forbidden gate firing zones for a particular thyristor in a 3-phase, 6-pulse converter. Specifically, it refers to Q1 in Fig. 21-50. The other thyristors have similar firing zones, but they occur at different times.



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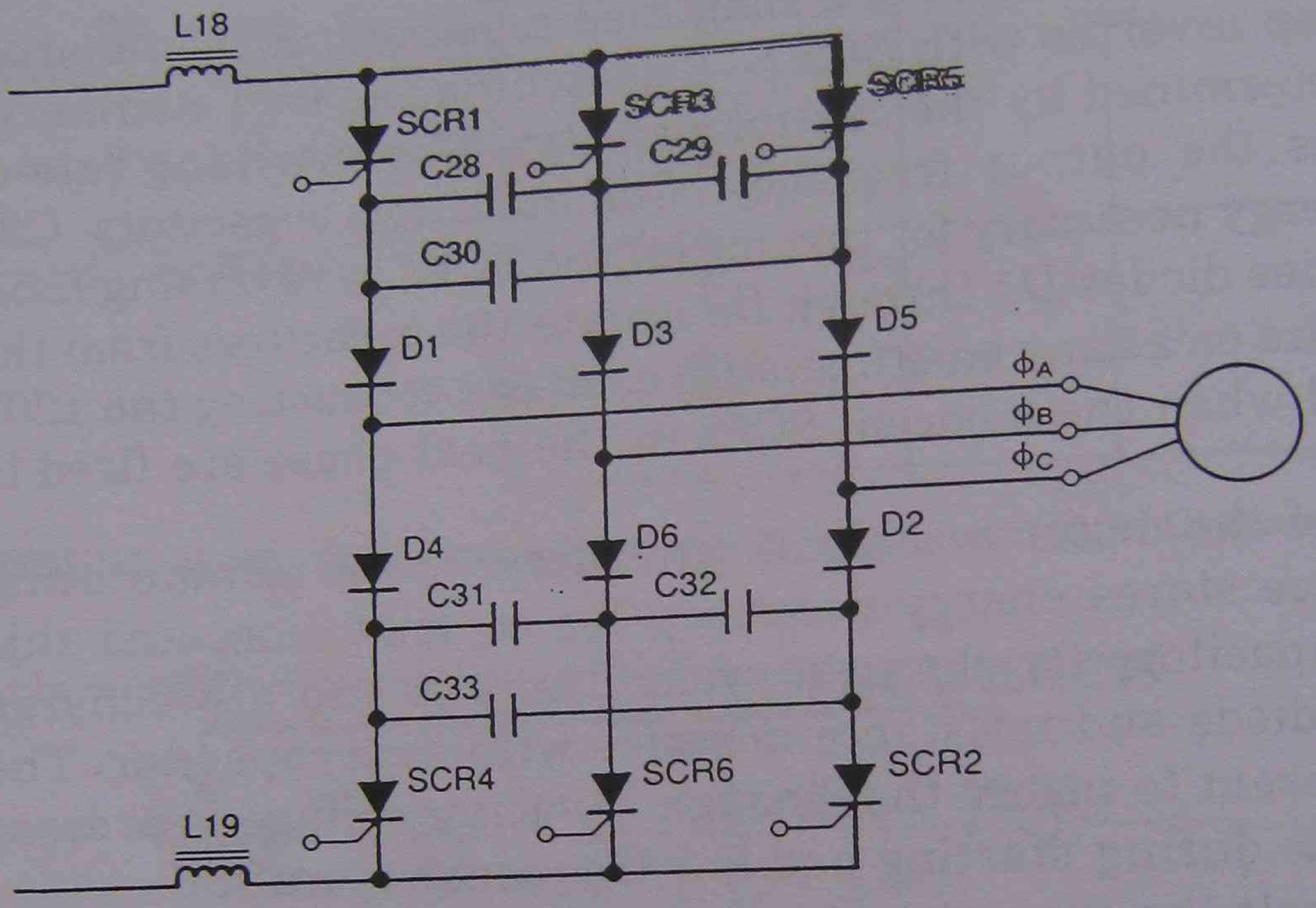


FIGURE 11–23 Inverter section, Graham CSI controller

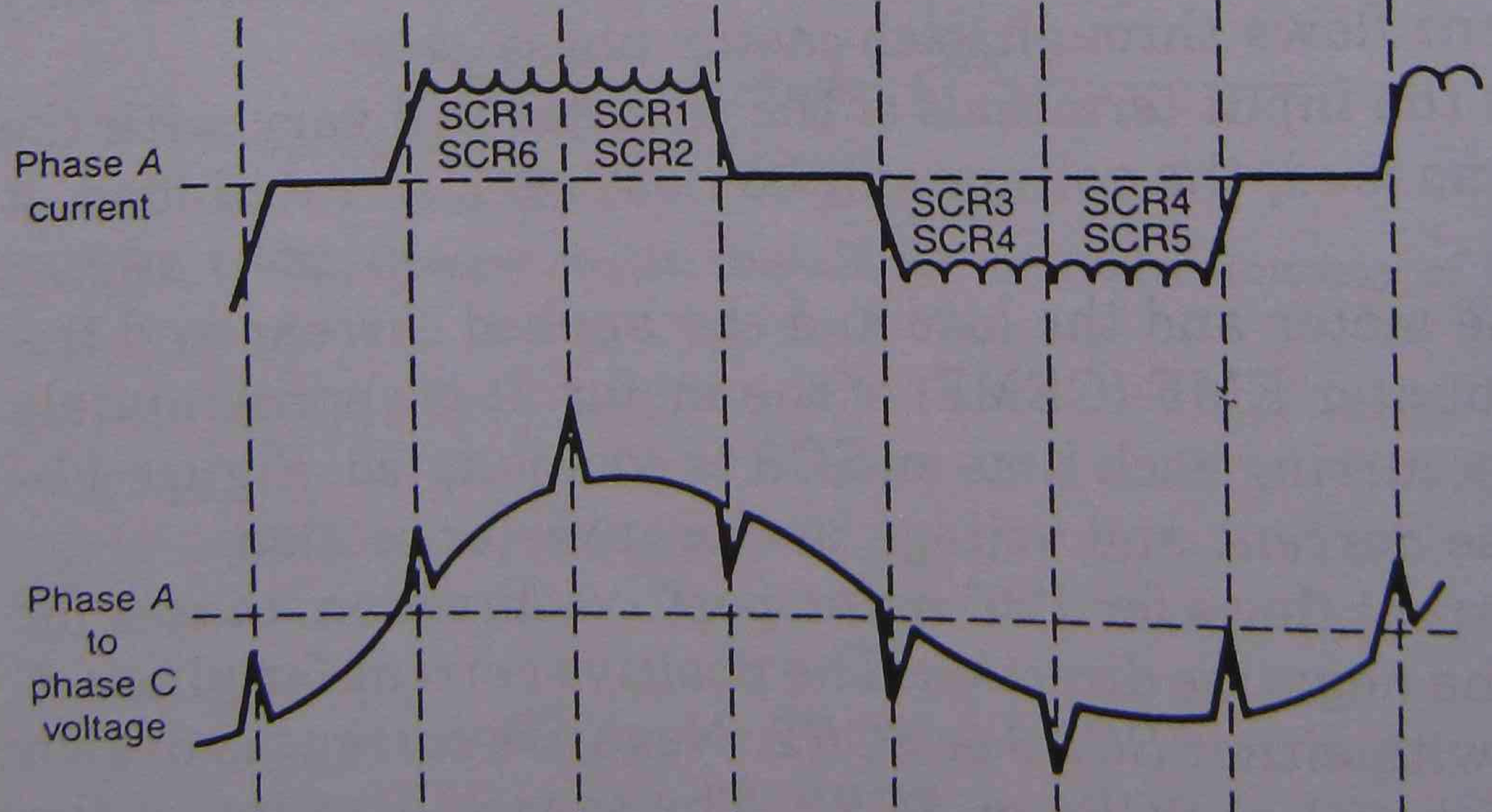
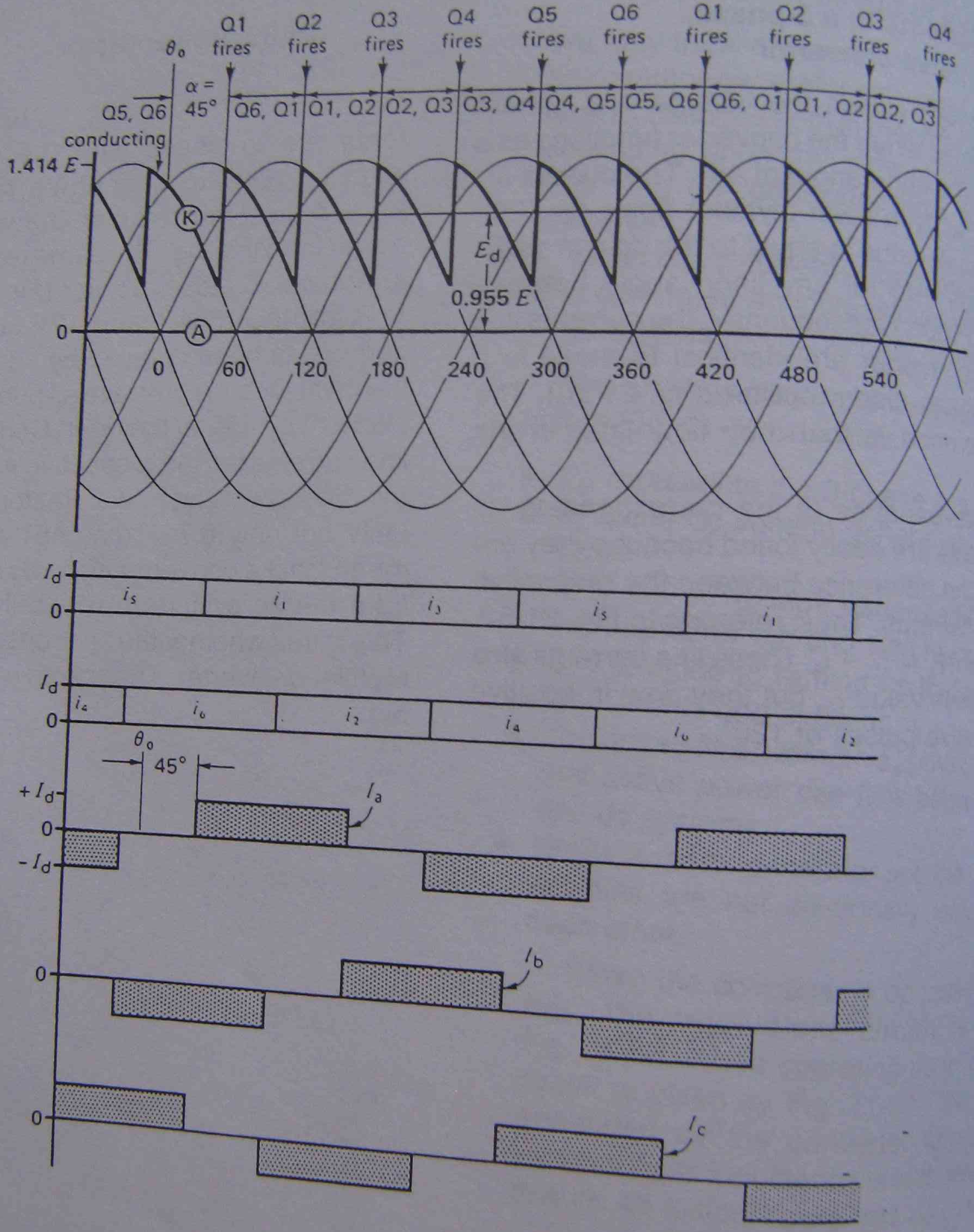
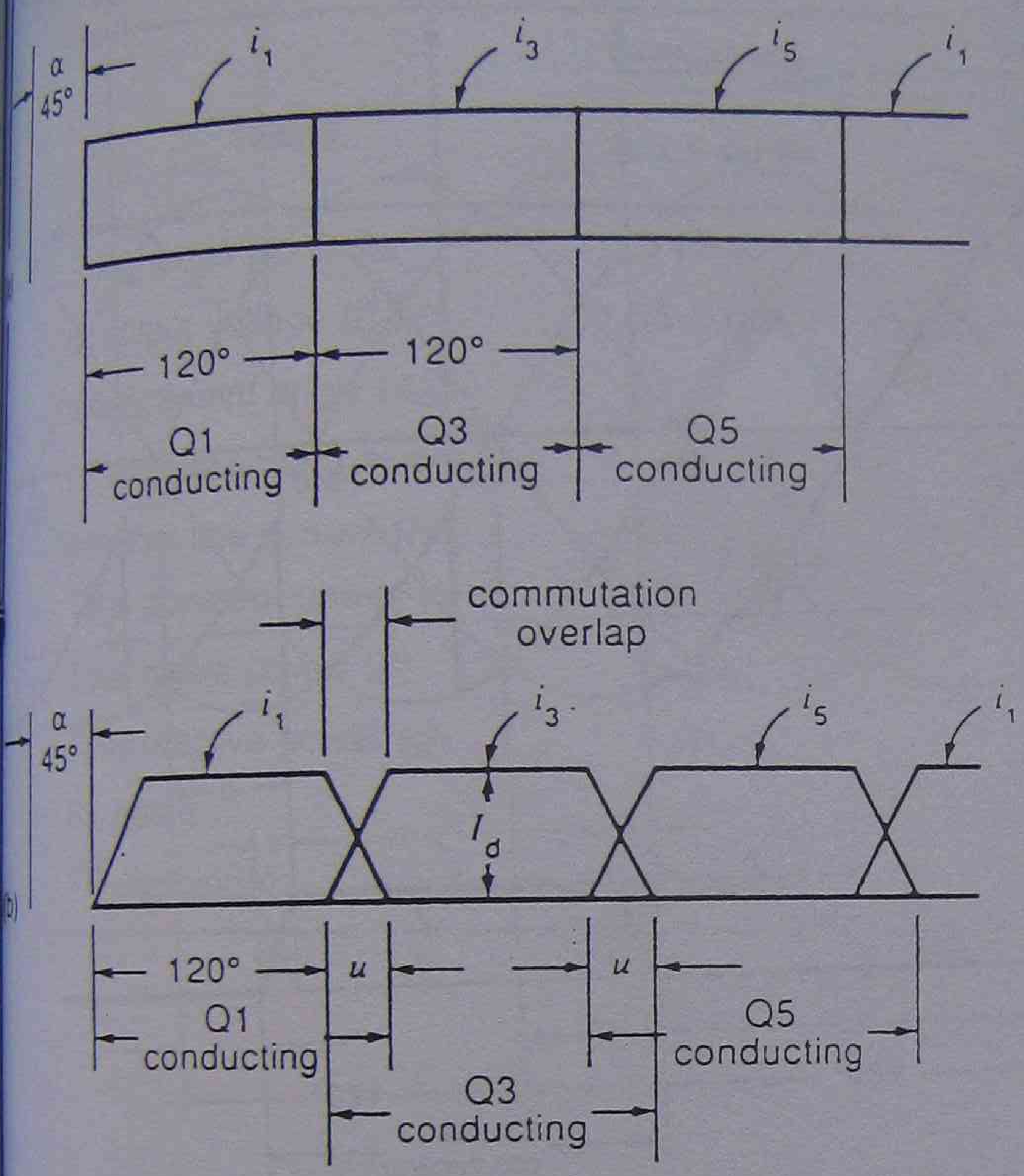


FIGURE 11–24 Synchrogram showing currents and voltages in inverter (courtesy of Graham Co.)



gure 21-58
ltage and current waveforms in the converter of Figure 21-50 with a delay angle of 45°.



gure 21-60
Instantaneous commutation in a rectifier when α = 45° (see Fig. 21-58).
Same conditions with commutation overlap of 30°, showing current waveshapes in Q1, Q3, Q5.

21.34 Commutation overlap

We mentioned in Section 21-9,. that the current in a three-pulse rectifier cannot switch instantaneously from one diode to the next. The commutation process takes time and this is also true for thyristors. Thus, in a six-pulse converter, the commutation from Q1 to Q3 to Q5 is not instantaneous (as shown in Figure 21-58), but is more like that shown in Figure 21-60b.

The transfer of I_d from one thyristor to the next is effected during the so-called commutation overlap period, defined by angle u. The amount of overlap varies with the current I_d. At full load, u lies typically between 20° and 30°. At light load it can be as small at 5°.

On account of commutation overlap, the current in each thyristor flows for a period of $120 + u$ degrees instead of 120°, as we have assumed so far. The commutation overlap modifies the waveshape of E_{AK}, but we do not have to examine this aspect of converter behavior.

The commutation overlap delays the current buildup by angle u. It also delays the current cutoff by the same angle. Owing to these delays, the effective firing angle is somewhat greater than α. This reduces the power factor of the converter in both the rectifier and inverter mode. It also reduces the average dc voltage E_{dc}.

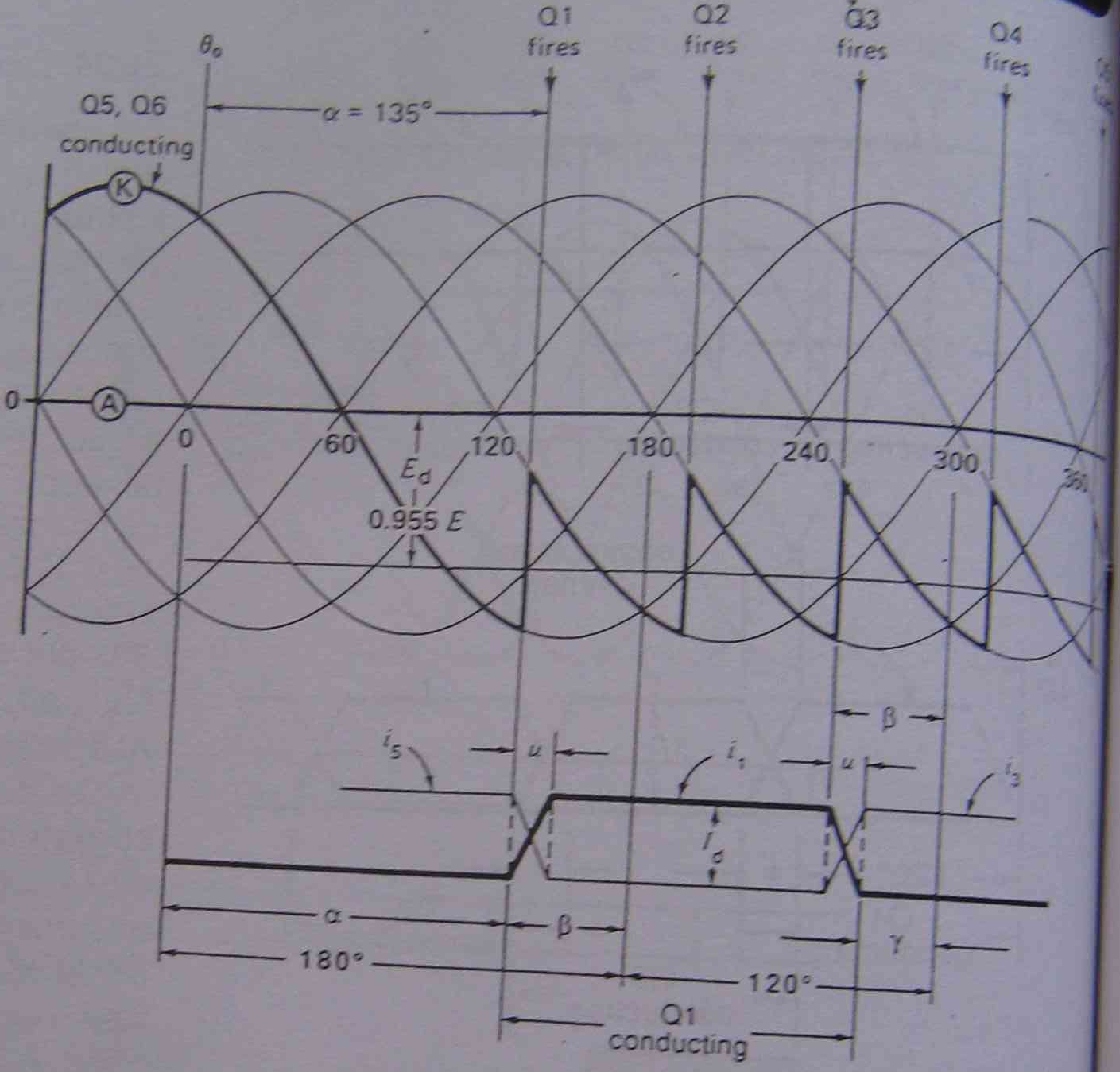


Figure 21-61
Waveshape of i_1 in thyristor Q1 for a delay angle α. The extinction angle γ permits Q1 to establish its blocking ability before the critical angle of 300° is reached. At 300° the anode of Q1 becomes positive with respect to its cathode. The figure also shows the relationship between angles α, β, γ and *u*.

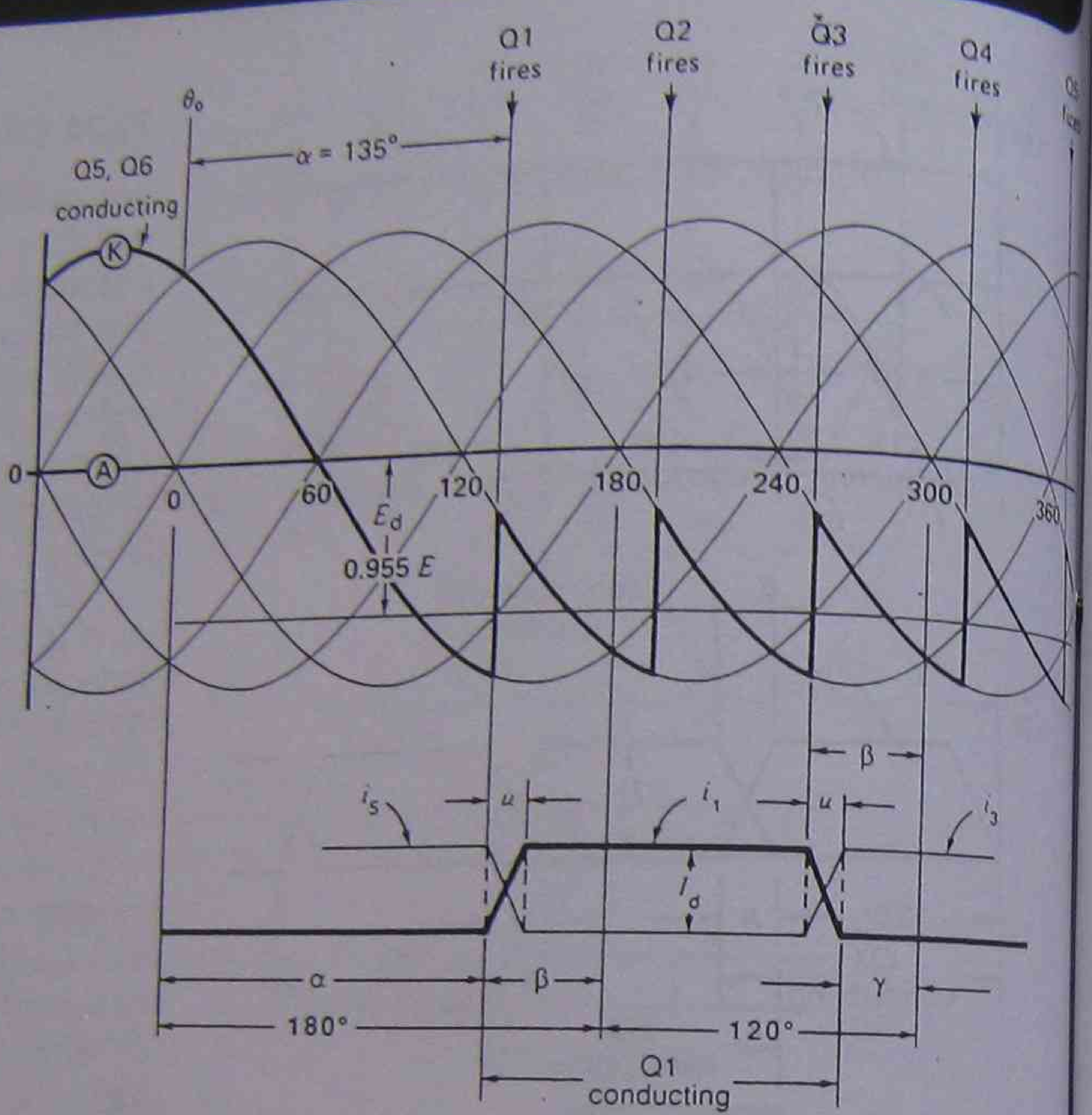
21.35 Extinction angle

We have seen that when a converter operates in the inverter mode, it is very important tha conduction be initiated prior to α = 180° Because the current in an ideal inverter flows for 120°, the conduction must also cease before the angle of (180 + 120) = 300° is reached. The interval between the end of commutation and 300° is called the *extinction angle* γ (Fig. 21-61). The extinction angle permits thyristor Q1 to recover its blocking ability before its anode (1) again becomes positive with respect to the cathode K. The value of γ lies typically between 15° and 20°.

In the case of an inverter, we often de the firing instant by the *angle of advanc* rather than by the angle of delay α. It is eas show that the following relationships e between the commutation angle *u*, the de angle α, the angle of advance β and th extinction angle γ:

$$\beta = 180 - \alpha \quad (21\text{-}1$$

$$\beta = u + \gamma \quad (21\text{-}2$$



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21.35 Extinction angle

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$$\beta = 180 - \alpha \qquad (21\text{-}1$$

$$\beta = u + \gamma \qquad (21\text{-}2$$

TABLE 22A PROPERTIES OF SOME RECTIFIER CONVERTERS (RESISTIVE LOAD)

ITEMS	Converter A 3ph, 6-pulse	Converter B 3 ph, 6-pulse + free-wheel diode	Converter C half bridge
Firing angle (α) limits	0 to 90°	60° to 120°	60° to 180°
dc output voltage (E_d)	$1.35\ E \cos \alpha$	$1.35E\ (1 - \cos(120 - \alpha))$	$0.675\ E\ (1 + \cos \alpha)$
displacement angle (ϕ_d)	α	$30 + \alpha/2$	$\alpha/2$
PF (displacement) = $\cos \phi_d$	$\cos \alpha$	$\cos (30 + \alpha/2)$	$\cos \alpha/2$
effective line current (I)	$0.816\ I_d$	$I_d \sqrt{(120 - \alpha)/90}$	$I_d \sqrt{(180 - \alpha)/180}$
Total apparent power (S)	$EI \sqrt{3}$	$EI \sqrt{3}$	$EI \sqrt{3}$
Total active power (P)	$E_d I_d$	$E_d I_d$	$E_d I_d$
Total reactive power (Q)	$P \tan \phi_d$	$P \tan \phi_d$	$P \tan \phi_d$
PF (total)	P/S	P/S	P/S
PF (distortion)	$P/S \cos \phi_d$	$P/S \cos \phi_d$	$P/S \cos \phi_d$

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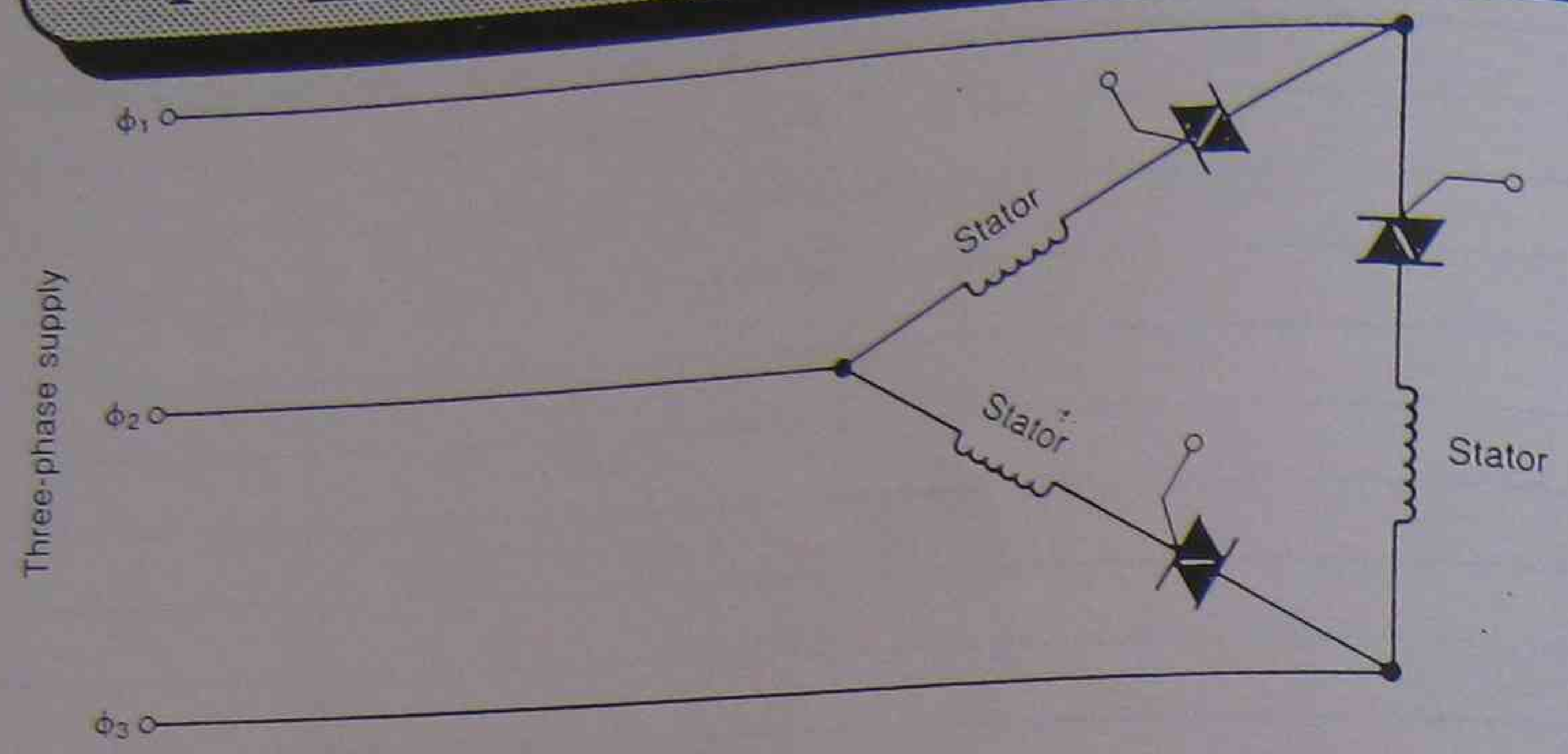
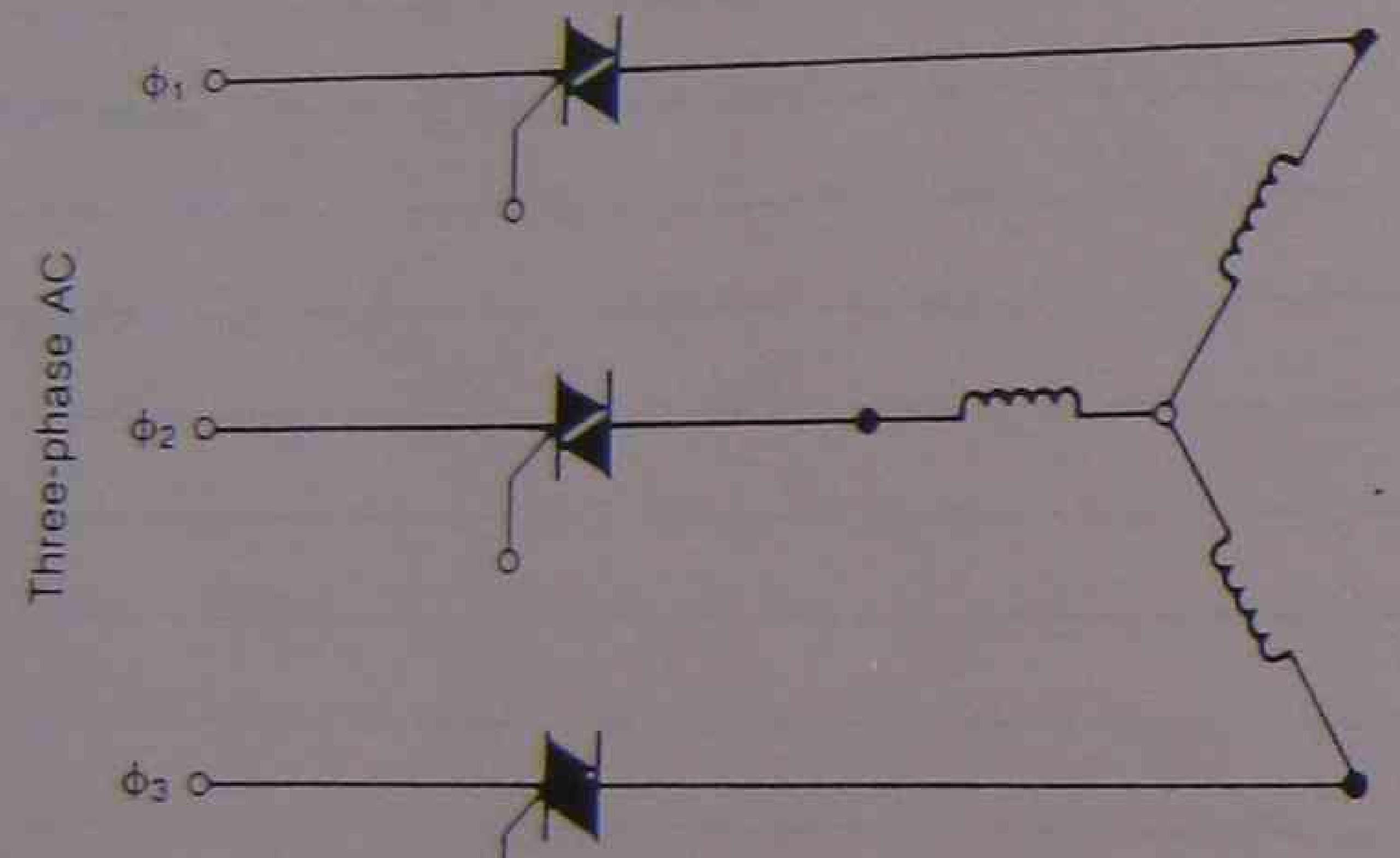
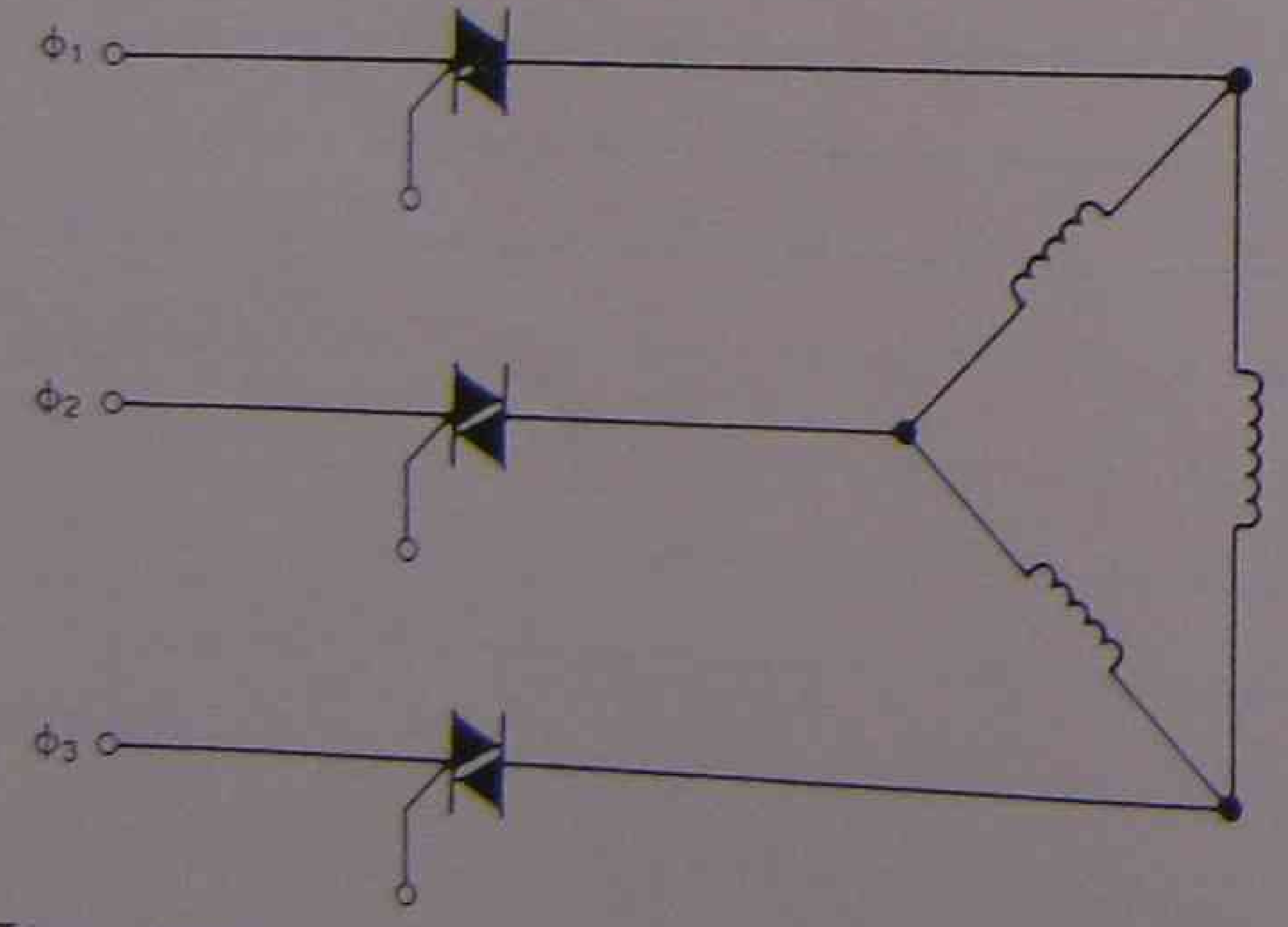


FIGURE 11–4 Triacs controlling three-phase motor

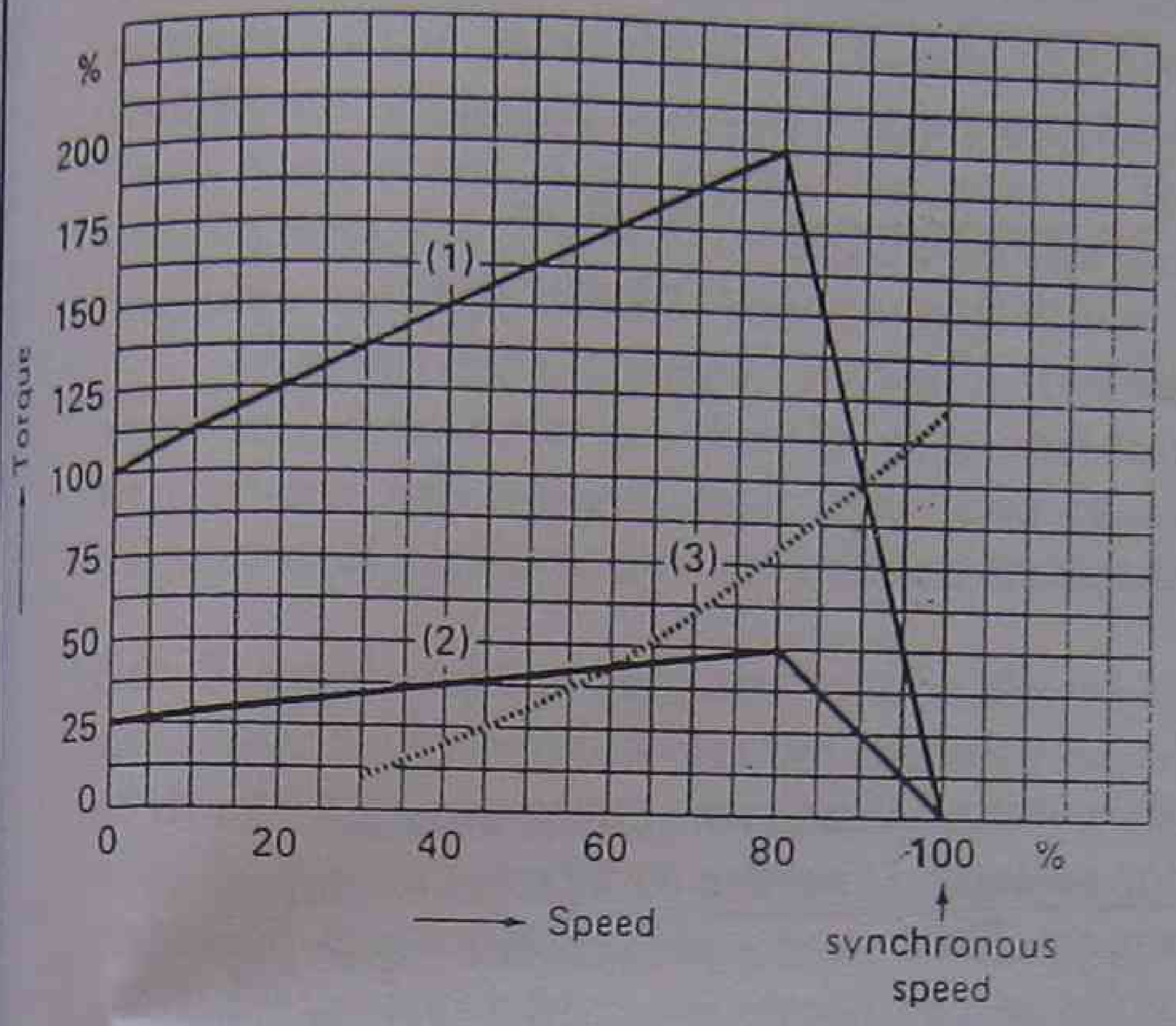


a. Triacs controlling, connected stators –



b. Triacs controlling, connected stators –

FIGURE 11–5 Triacs controlling three-phase motors



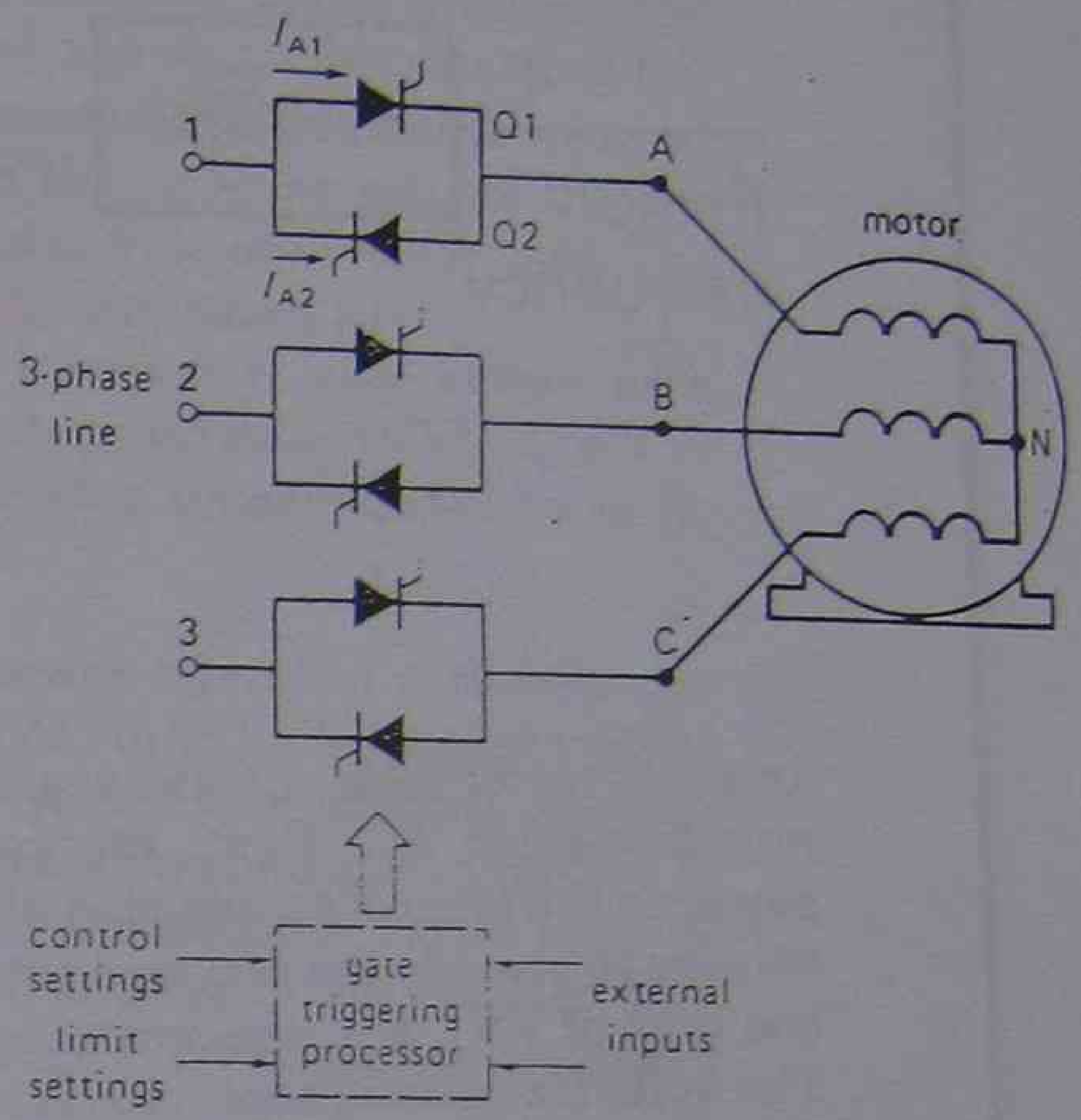


Figure 23-6
Variable-voltage speed control of a squirrel-cage induction motor.

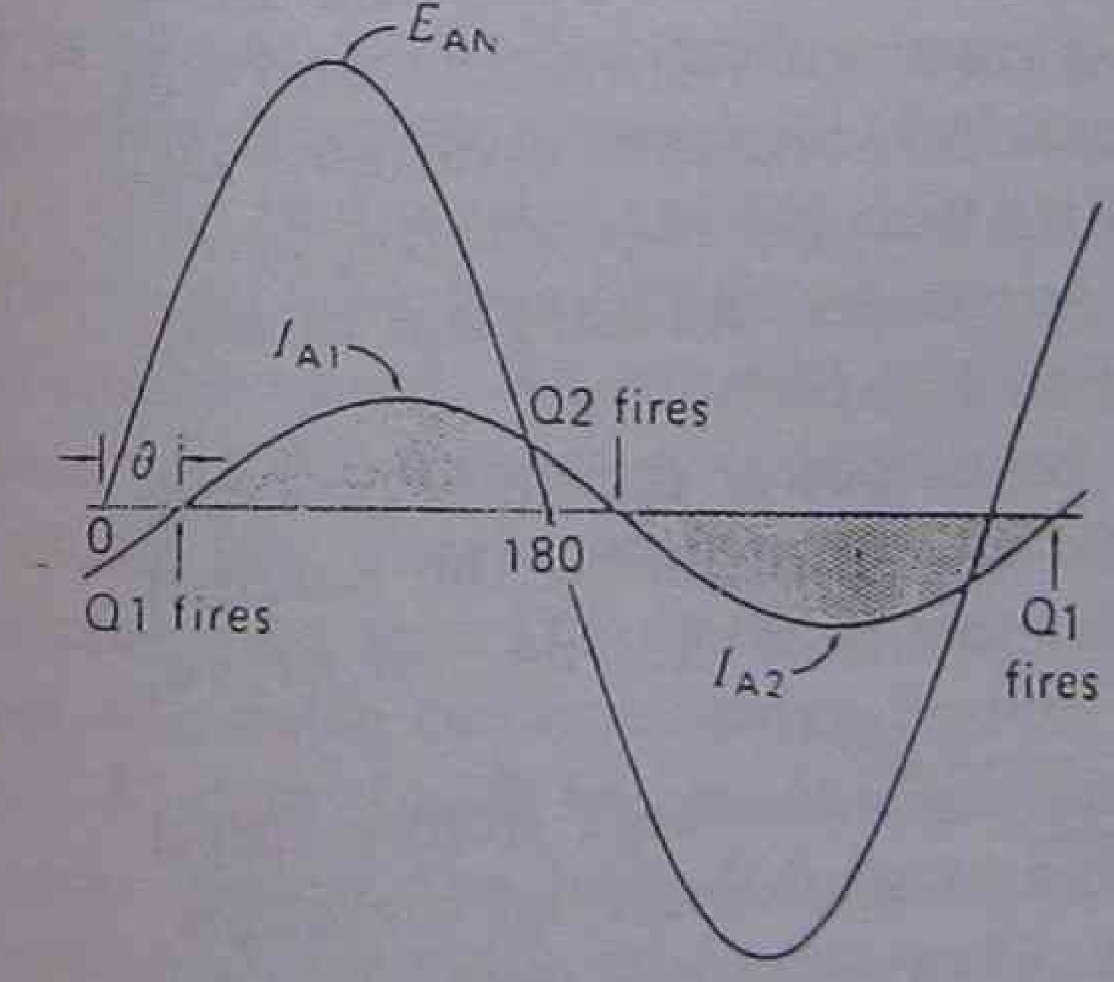


Figure 23-7
Waveshapes at rated voltage.

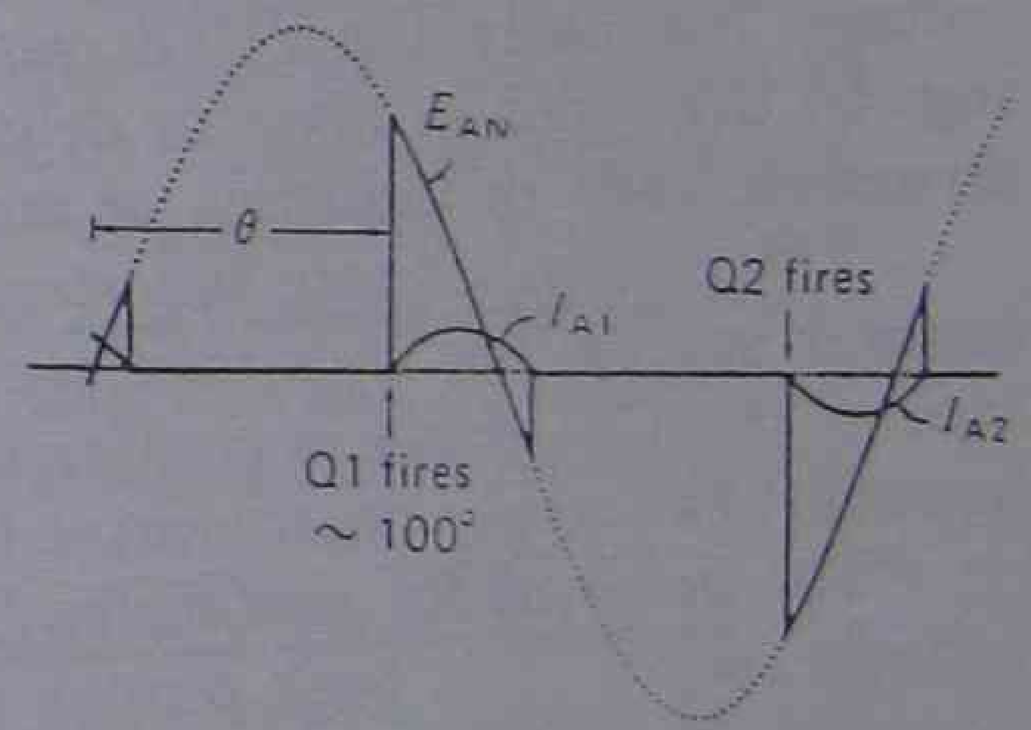
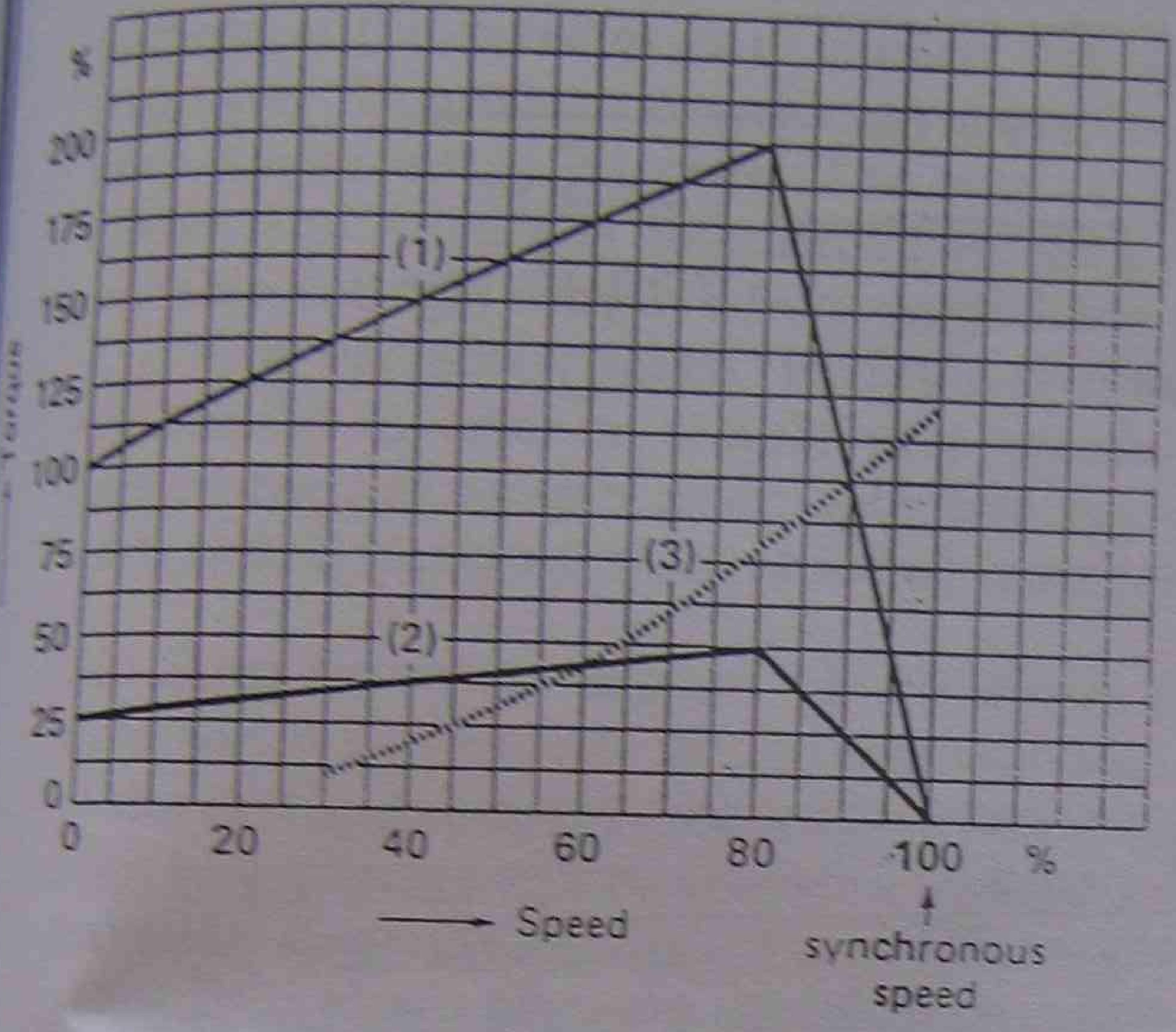


Figure 23-8
Waveshapes (very approximately) at 50% rated voltage



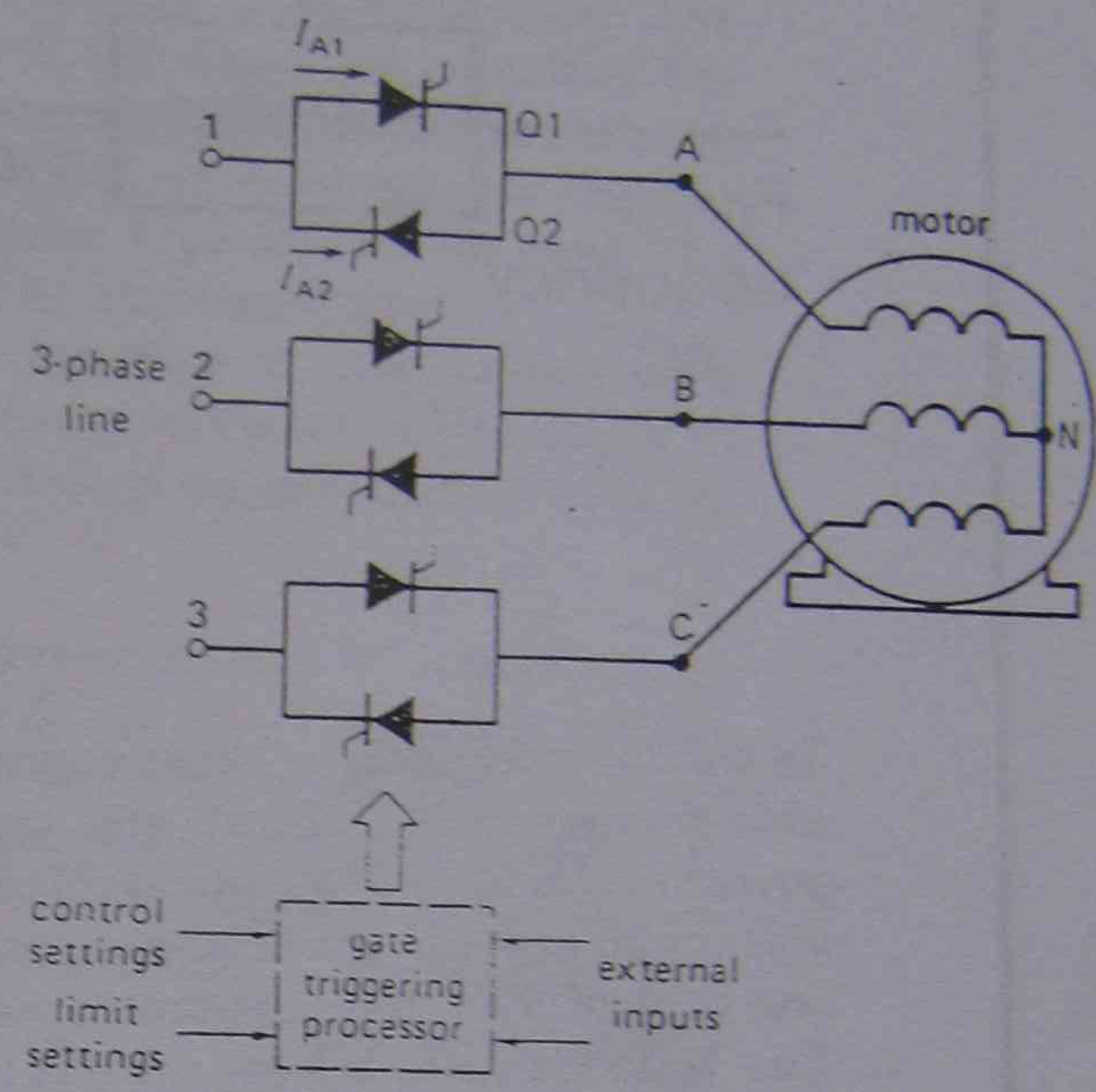
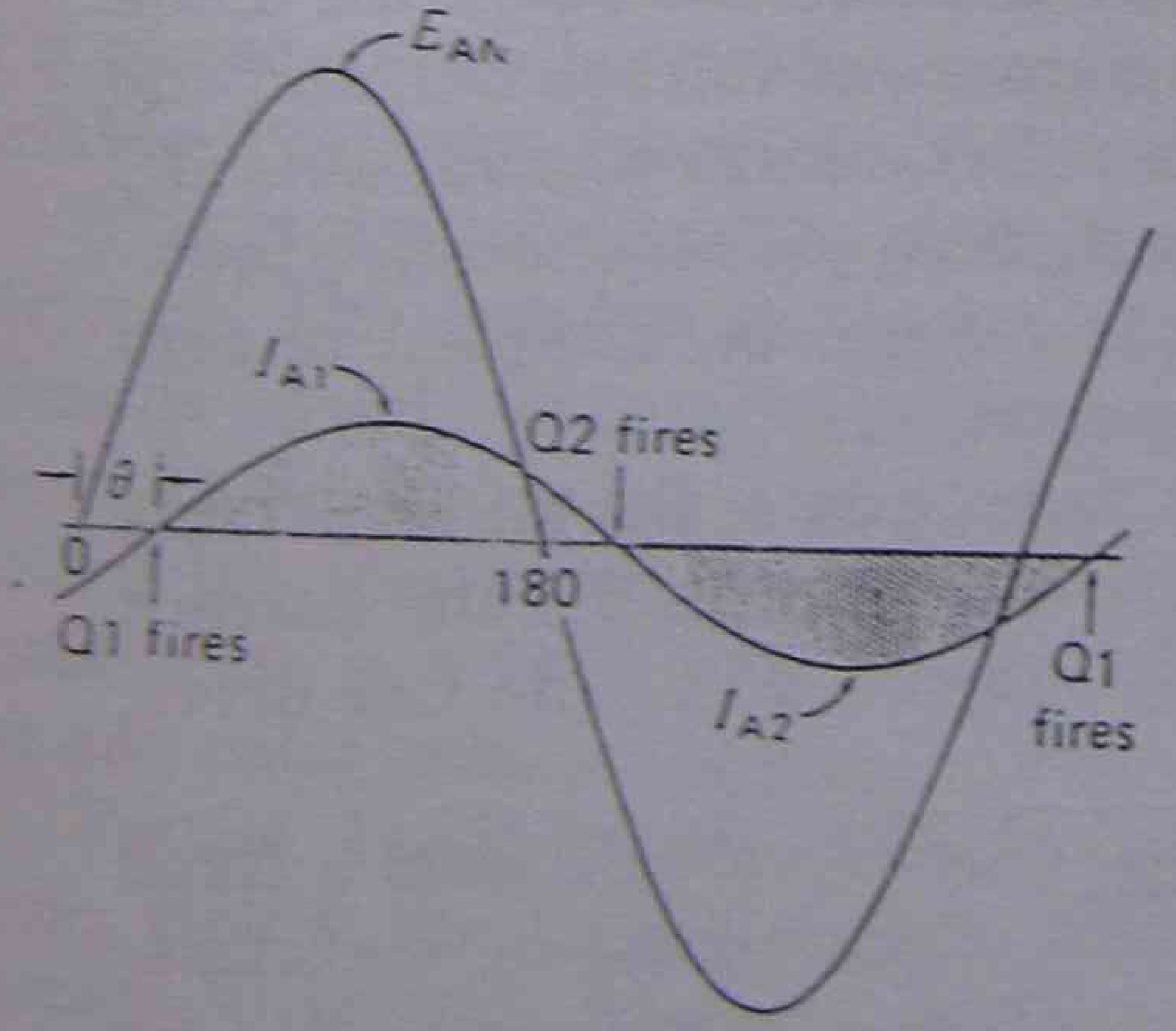


Figure 23-6
Variable-voltage speed control of a squirrel-cage induction motor.



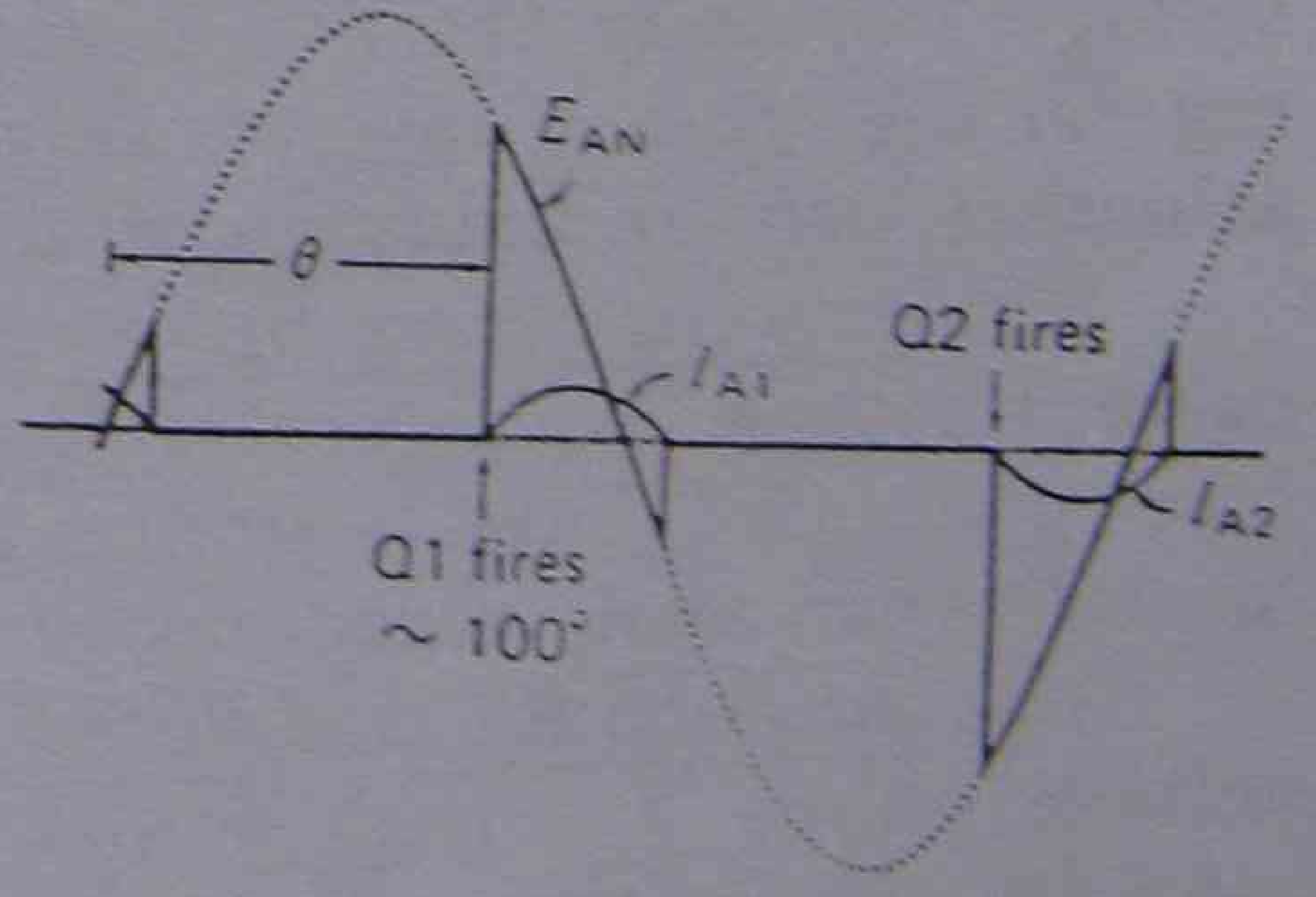
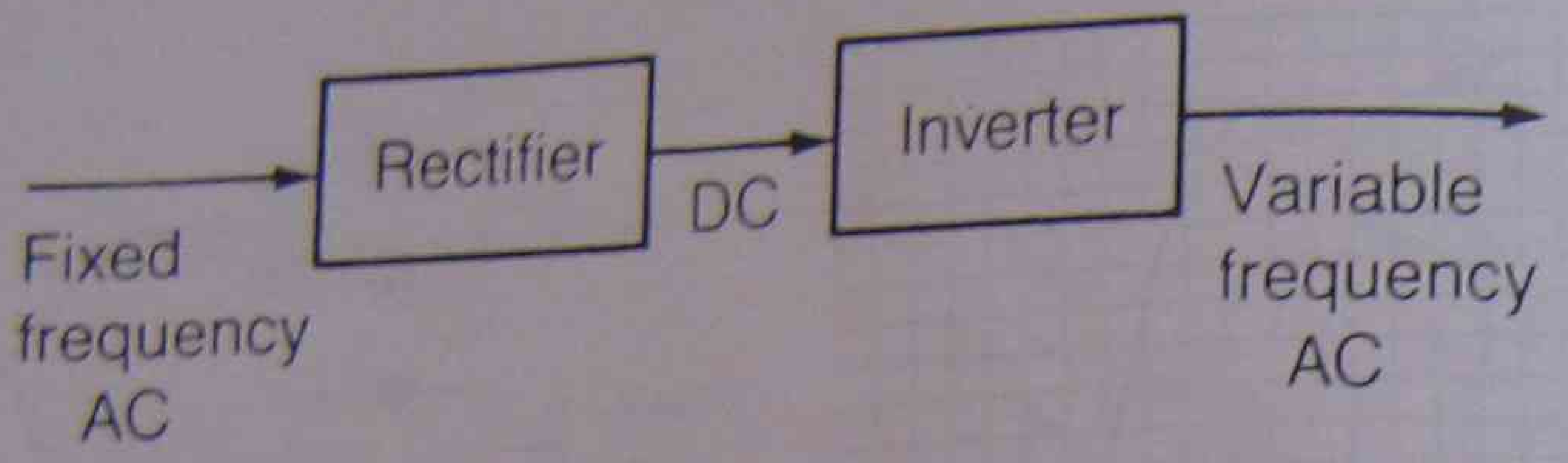


Figure 23-8
Waveshapes (very approximately) at 50% rated voltage.

Figure 23-7
Waveshapes at rated voltage.



In the inverter section, three approaches are used to convert DC to adjustable-frequency AC. First, the *variable-voltage inverter (VVI)* takes input power in the form of an adjustable DC source. This source presets the input DC voltage to provide the required output voltage amplitude from the inverter. In one type of VVI, a phase-controlled bridge rectifies the incoming AC voltage. The volts/Hz ratio is kept constant by changing the amplitude of the rectified DC as the frequency is changed. A second type of VVI replaces the phase-controlled rectifier bridge with a diode bridge and a DC regulator or chopper. This system, therefore, has a rectifier that is divided into two parts: the diode bridge, which converts fixed frequency AC to a constant voltage DC, and the regulator or chopper, which changes the constant DC voltage to a variable DC voltage. Normally, the VVI drives lack the ability to apply regenerative braking. Of the types of inverter drives, the VVI drives are the simplest in construction, used in industry for applications up to 400 HP.

The *current-source inverter (CSI)* takes input power for an adjustable current source, not a voltage source, as in the VVI. Except for the current source, the CSI drive is similar in construction to the VVI. The CSI drive, however, can apply regenerative braking to a motor.

The *pulse-width modulated (PWM) inverter* takes voltage from a fixed voltage source. The peak output voltage applied to the motor is, therefore, constant. The average value of the output voltage wave form is controlled by changing the width of the zero voltage interval in the output wave form.

11–2.3 Variable-Voltage, Variable-Frequency Control

This type of induction motor speed control uses a circuit called an inverter. The purpose of the inverter is to control the speed of the motor by adjusting the frequency. To produce a constant torque for the motor, the inverter drive must keep a constant V/Hz ratio. The way in which the inverter adjusts the frequency and voltage is determined by the particular type of inverter used. The variable-voltage, variable-frequency drive will be discussed first.

The variable-voltage, variable-frequency inverter is also known as a voltage-fed inverter or, simply, as a variable-voltage inverter (VVI). The VVI can be further broken down into two types: six-step (quasi-square wave inverter) and pulse-width modulated inverters.

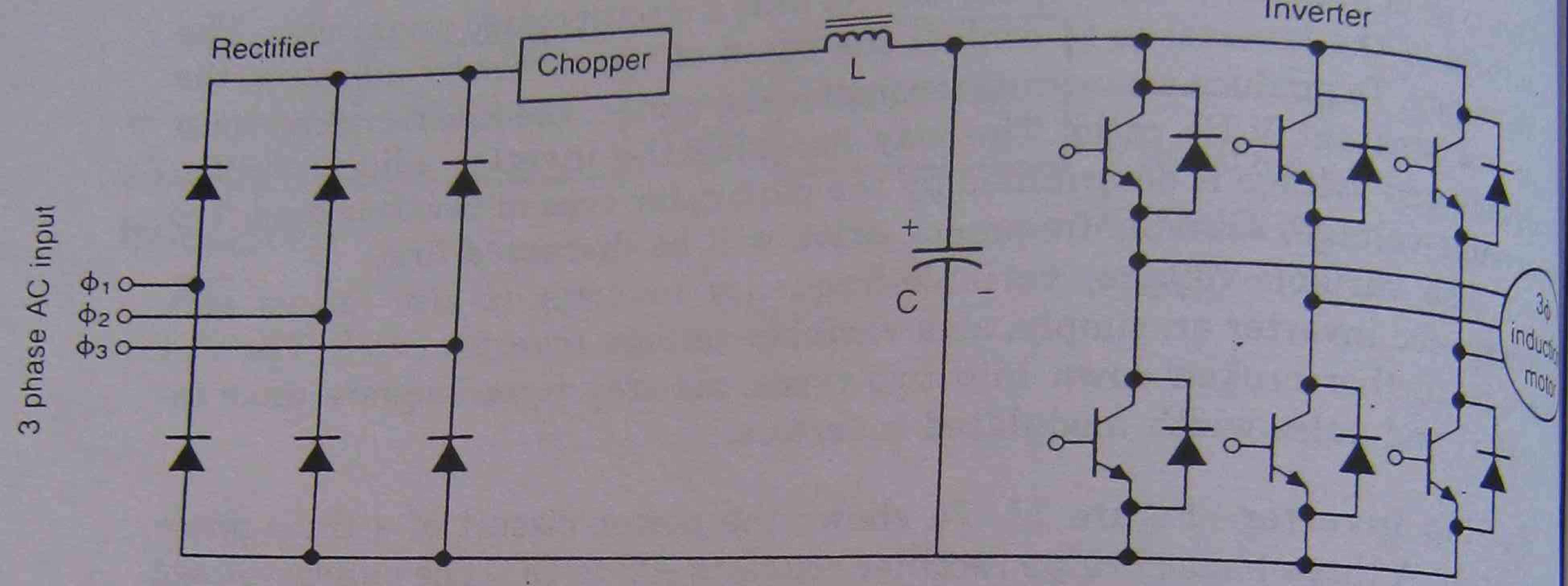
Six-Step Inverter Figure 11–7a shows the power circuit of a three-phase inverter. A three-phase bridge rectifier converts AC to DC. The output voltage of the rectifier section is varied by a DC chopper. A thyristor chopper is preferable to a transistor chopper, which must use several transistors connected in parallel. Regardless of the type of chopper used, the chopper varies the constant DC voltage from the rectifier, which is then applied to the inverter. This type of inverter is called voltage fed because a large filter capacitor provides a stiff voltage supply to the inverter.

The inverter output voltage wave forms are not affected by the nature of the load. Figure 11–7b shows another way to vary the input voltage. In this method, the uncontrolled diode rectifier and the chopper regulator are replaced by a phase-controlled bridge rectifier. The principle of the variable-voltage, variable-frequency speed control method is shown in Figures 11–8 and 11–9.

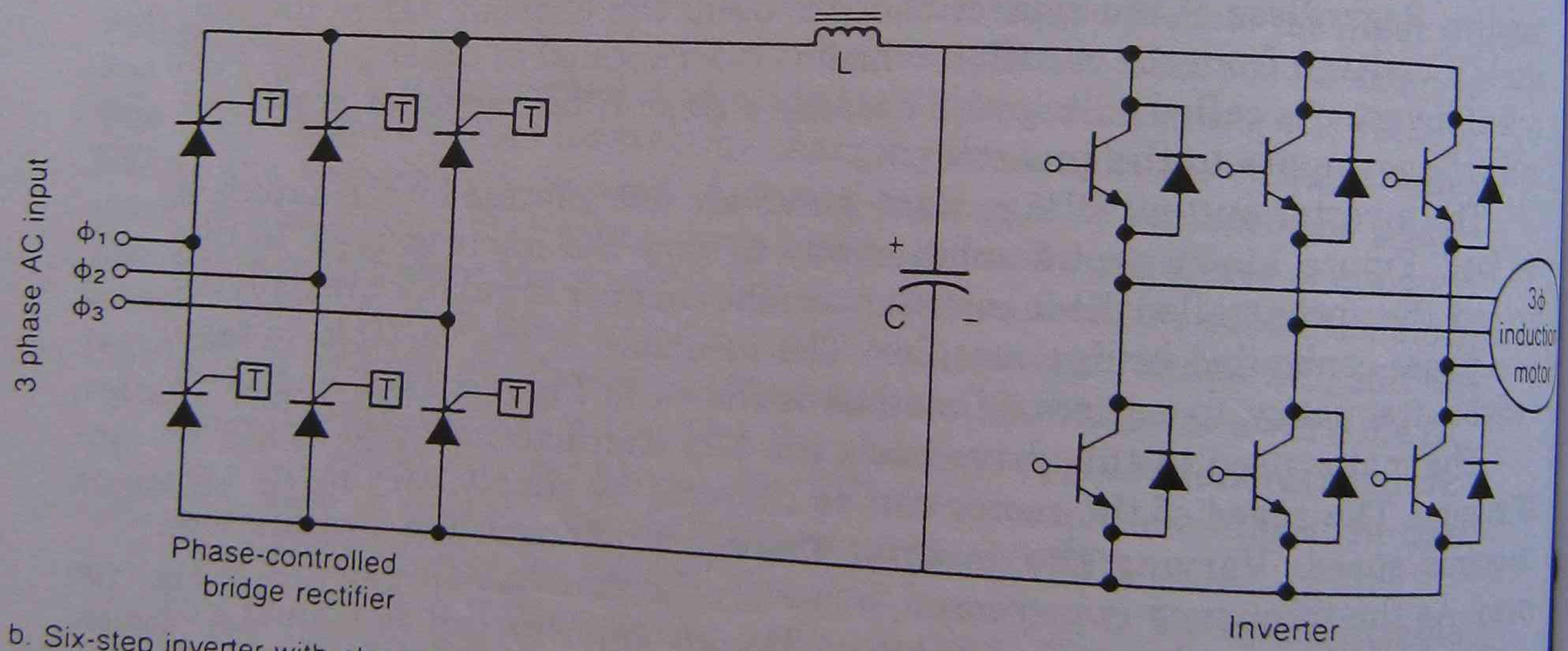
The motor used in this drive has a low slip characteristic that improves efficiency. The speed of the motor can be changed by simply varying its synchronous speed. Varying the inverter frequency changes the synchronous speed. As the frequency is increased, however, the machine air gap *flux* falls, causing low developed torque capability. The air gap flux can be maintained constant, as in a DC shunt motor, if the voltage is varied with frequency so that the ratio remains constant.

Figure 11–8 shows the desired voltage-frequency relationship of the motor. Below the base frequency, the air gap flux is kept constant by the constant V/Hz ratio, which keeps the torque constant. At a very low frequency, the stator resistance is greater than the leakage inductance. To counter this effect, additional voltage is applied. At the base frequency, the input voltage regulator establishes full-motor voltage. Beyond this point, as frequency increases, the torque decreases because of loss of air gap flux. From this point on, the machine operates in a constant horsepower mode, as shown in Figure 11–9. In the constant horsepower mode, each torque-speed curve corresponds to a particular voltage and frequency combination at the machine terminal.

VVI



a. Six-step inverter with a conventional rectifier, chopper, and inverter



b. Six-step inverter with phase-controlled rectifier and inverter

FIGURE 11–7 Three-phase inverter

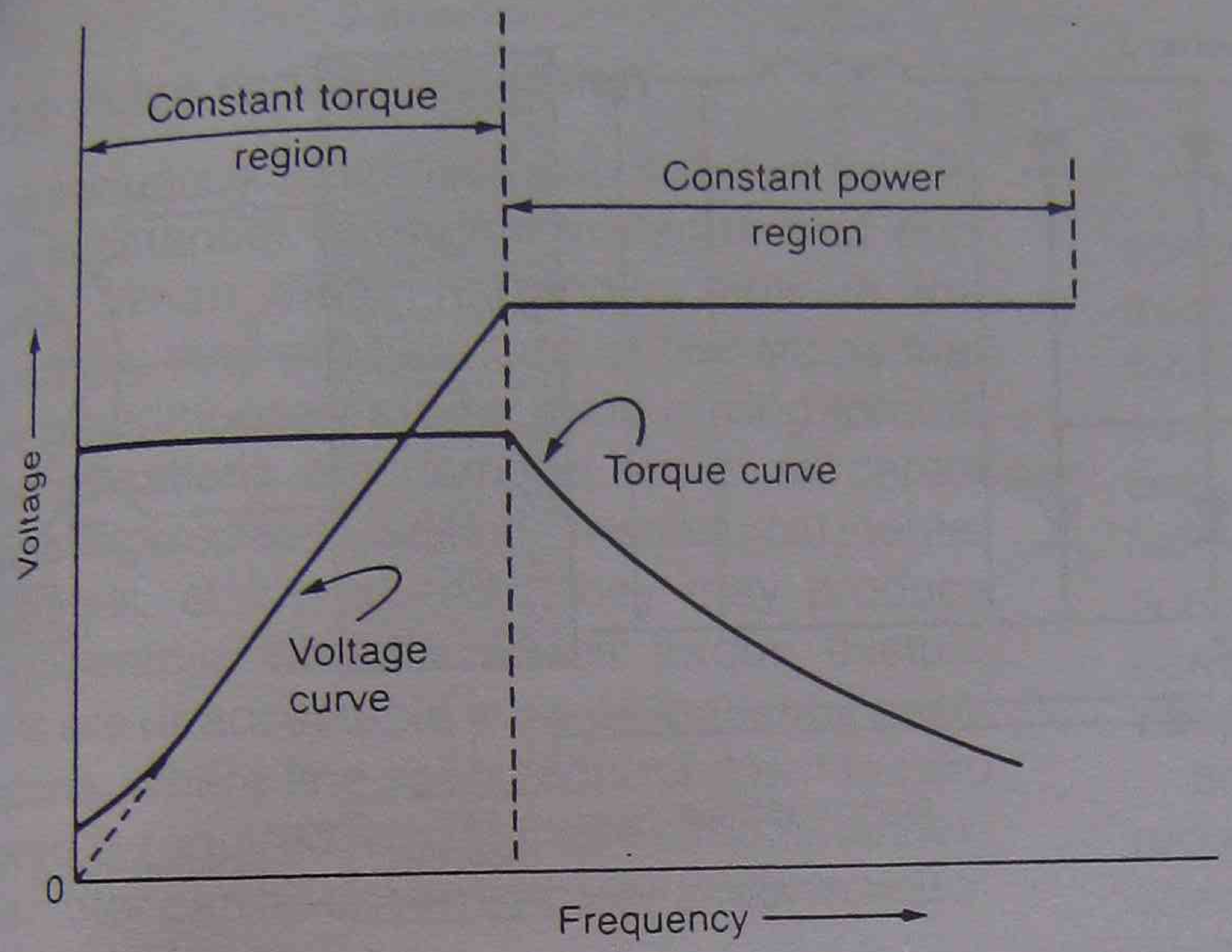


FIGURE 11–8 Voltage-frequency curve of induction motor

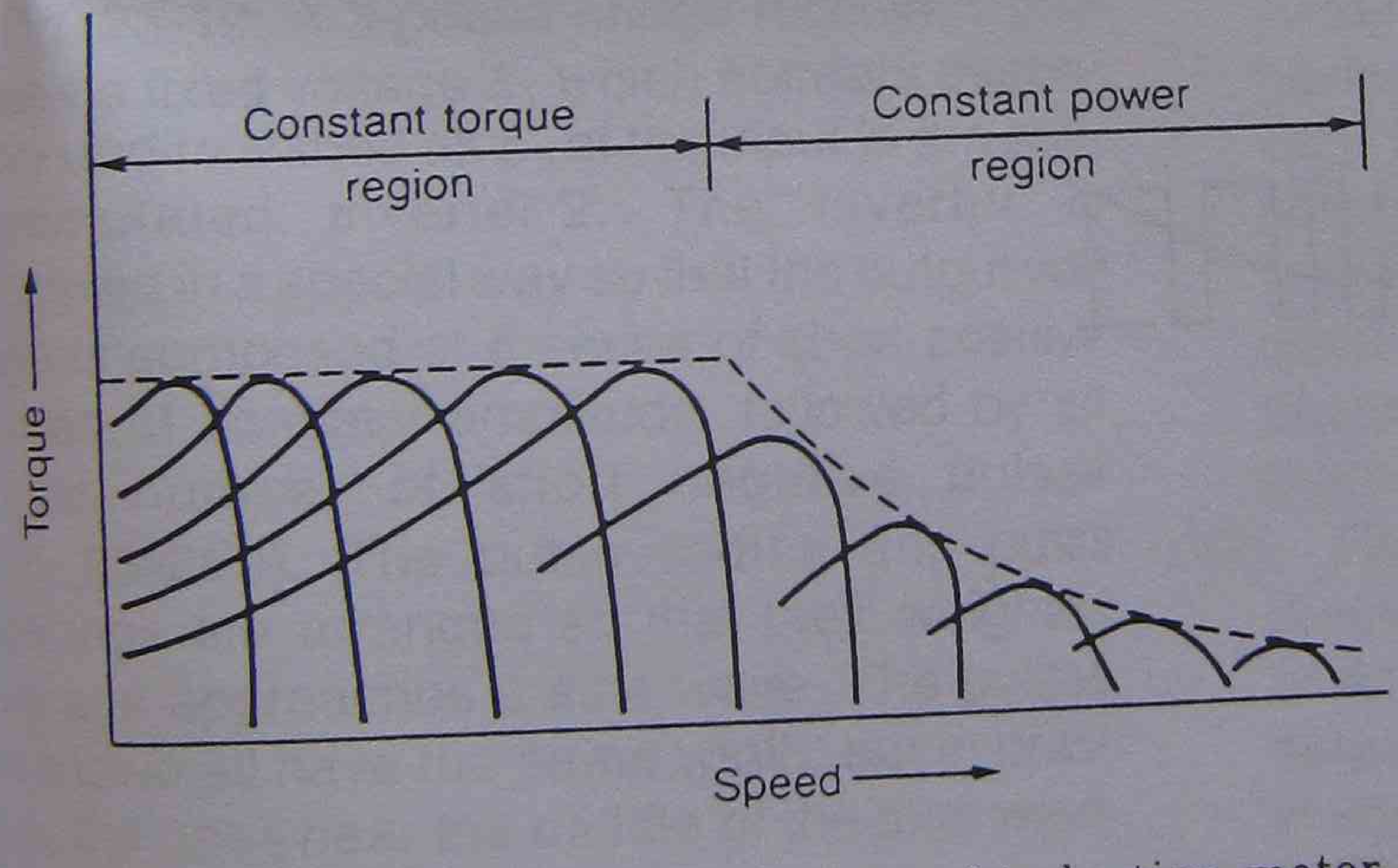


FIGURE 11–9 Torque-speed curves of induction motor with variable-voltage, variable-frequency power supply

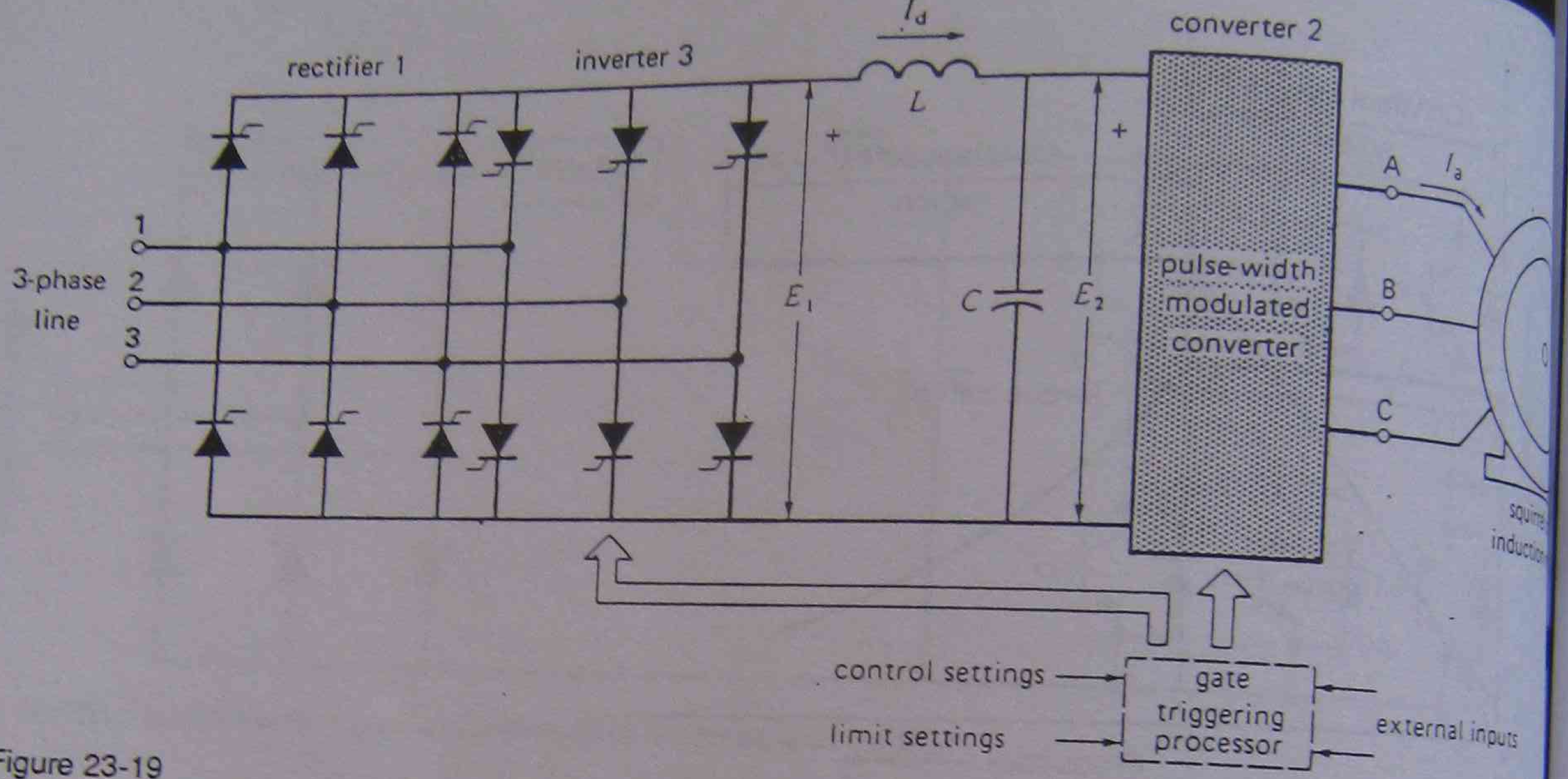


Figure 23-19
Speed control by pulse width modulation.

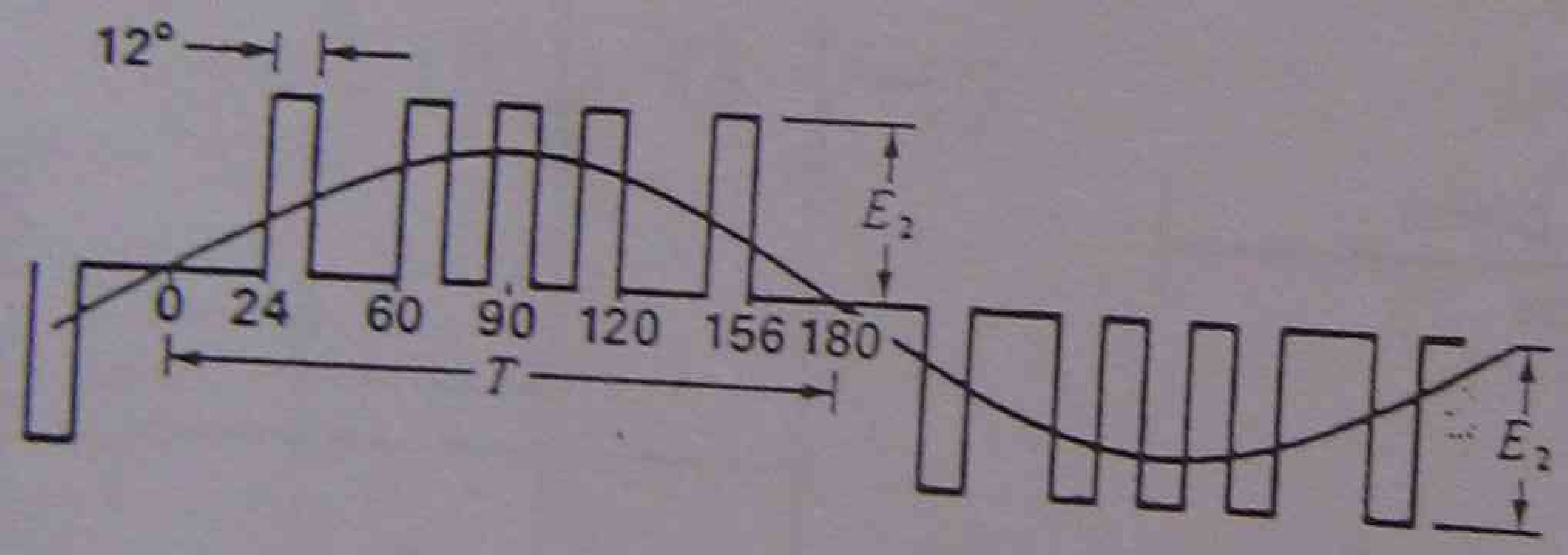


Figure 23-20a
Voltage waveform across one phase.

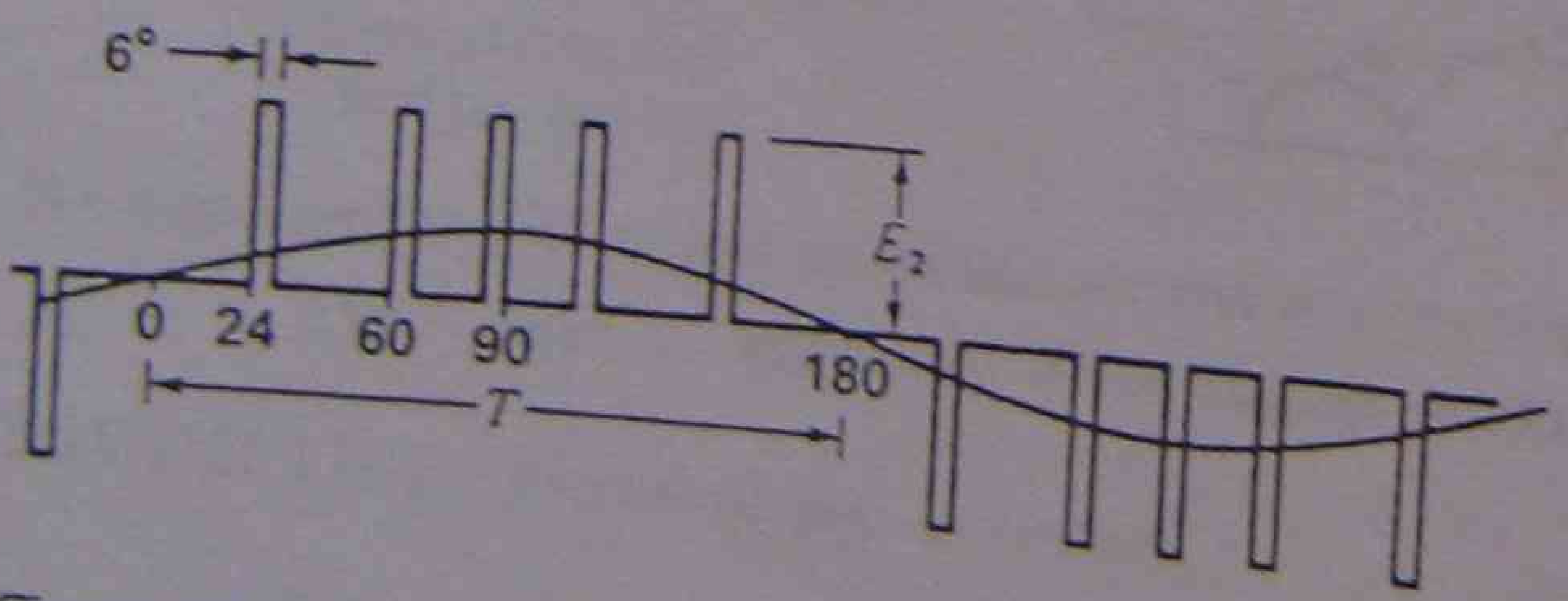


Figure 23-20b
Waveform yielding the same frequency but half the voltage.

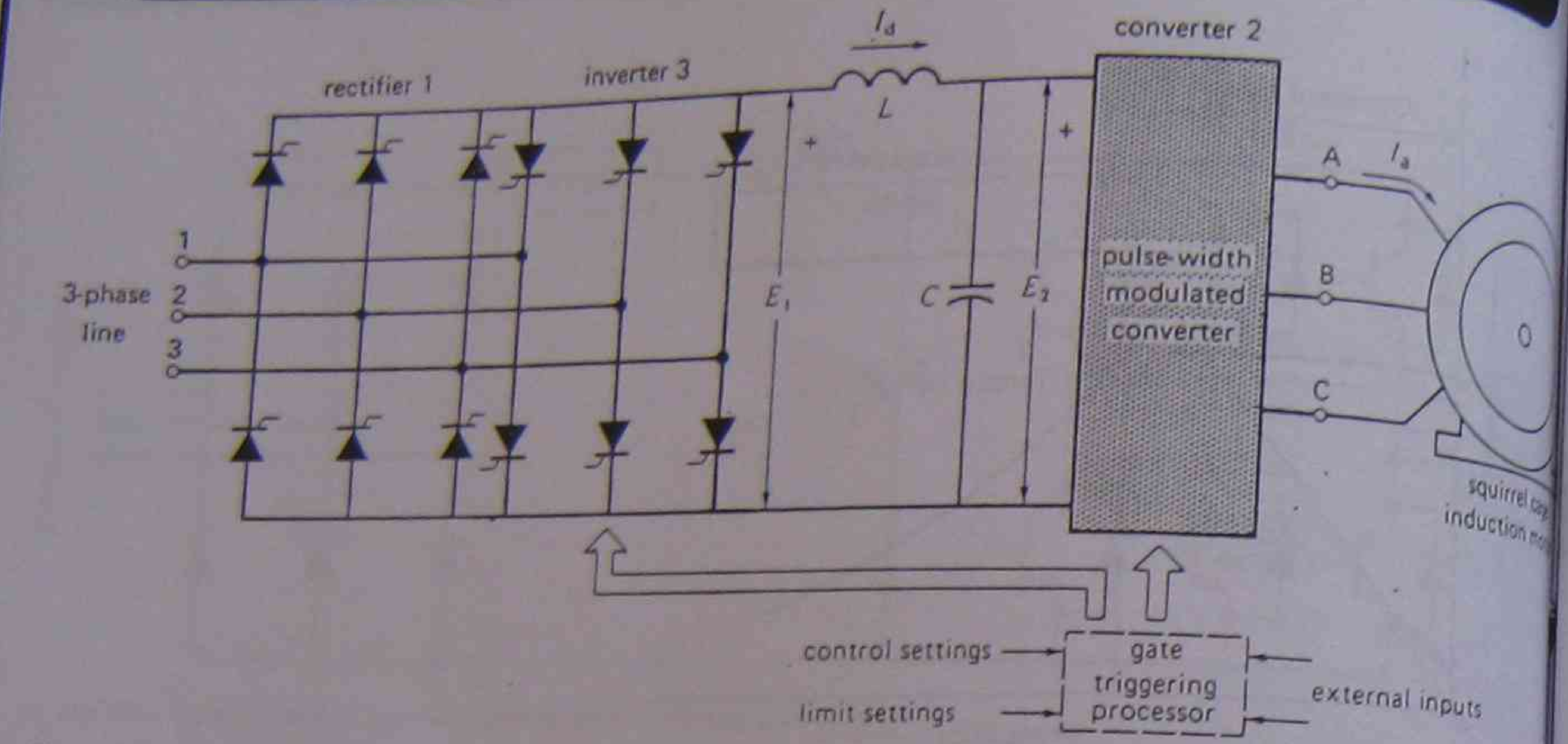


Figure 23-19
Speed control by pulse width modulation.

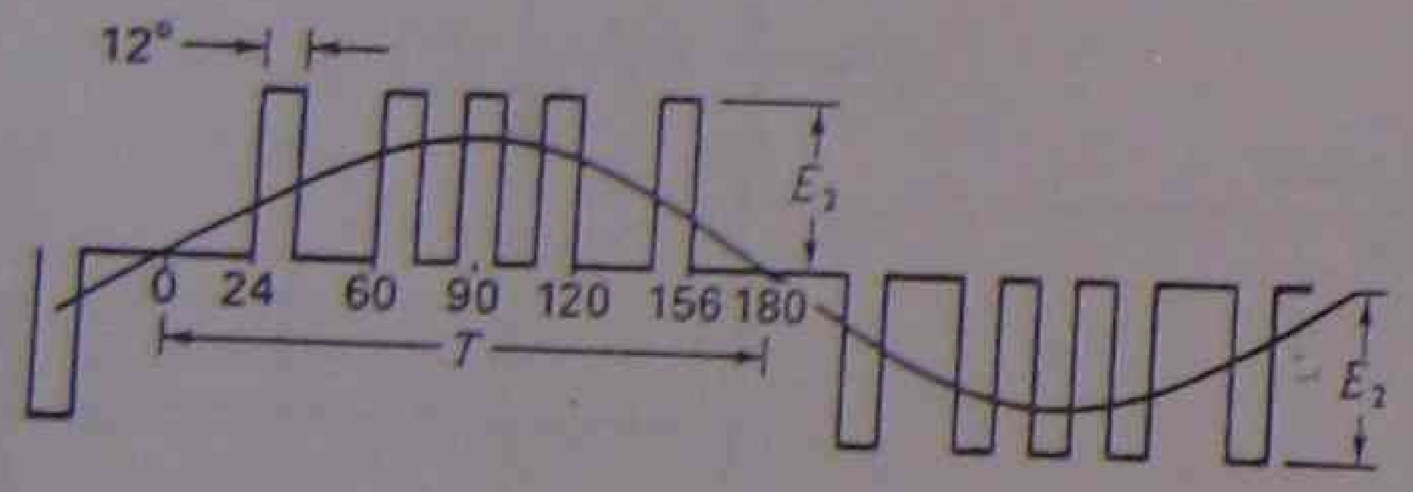


Figure 23-20a
Voltage waveform across one phase.

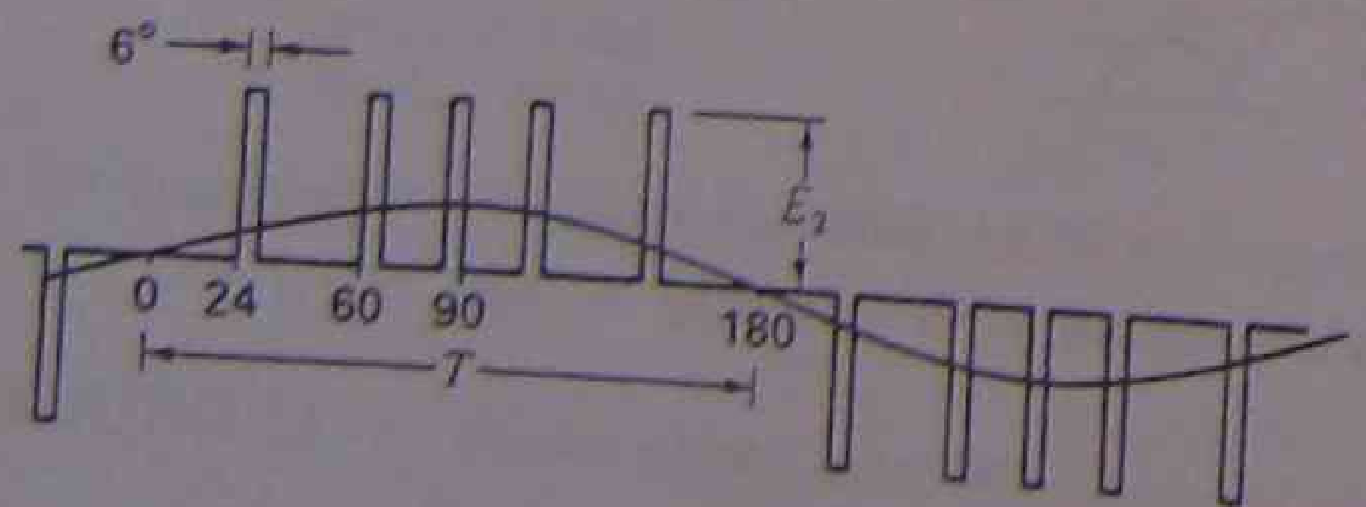


Figure 23-20b
Waveform yielding the same frequency but half the voltage.

23.9 Pulse width modulation

The frequency converters discussed so far create substantial harmonic voltages and currents. When these harmonics flow in the windings, they produce torque pulsations that are superimposed on the main driving torque. The pulsations are damped out at moderate and at high speeds owing to mechanical inertia. However, at low speeds, they may produce considerable vibration. Such torque fluctuations are unacceptable in some industrial applications, where fine speed control down to zero speed is required. Under these circumstances, the motor can be driven by *pulse width modulation* techniques.

To understand the technique, consider the voltage-fed frequency converter system shown in Fig. 23-19. A 3-phase bridge rectifier 1 produces a fixed voltage E_1 which appears essentially undiminished as E_2 at the input to the self-commutated inverter 2. The inverter is triggered in a special way so that the output voltage is composed of a series of short positive pulses of constant amplitude, followed by an equal number of short negative pulses (Fig. 23-20a). The pulse widths and pulse spacings are arranged so that their weighted average approaches a sine wave. The pulses as shown all have the same width, but in practice, the ones near the middle of the sine wave are made broader than those near the edges. By increasing the number of pulses per half cycle, we can make the output frequency as low as we please. Thus, to reduce the output frequency of Fig. 23-20a by a factor of 10, we increase the pulses per half-cycle from 5 to 50.

The pulse widths and pulse spacings are specially designed so as to eliminate the low-frequency voltage harmonics, such as the 3rd, 5th, and 7th harmonics. The higher harmonics, such as the 17th, 19th, etc., are unimportant because they are damped out, both mechanically and electrically. Such pulse width modulation produces output currents having very low harmonic distorsion. Consequently, torque vibrations at low speeds are greatly reduced.

In some cases, the output voltage has to be reduced while maintaining the same output frequency. This is done by reducing all the pulse widths in proportion to the desired reduction in output voltage. Thus, in Fig. 23-20b, the pulses are half as wide as in Fig. 23-20a, yielding an output voltage half as great, but having the same frequency. We can therefore vary both the output frequency and output voltage using a fixed dc input voltage. As a result, a simple diode bridge rectifier can be used to supply the fixed dc link voltage. The power factor of the 3-phase supply line is therefore high.

Regenerative braking can be achieved, but during such power reversal, current I_d reverses while the polarity of E_2 remains the same. Consequently, an extra inverter 3 has to be placed in reverse parallel with rectifier 1 in order to feed power back to the line (Fig. 23-19). Rectifier 1 is automatically blocked while inverter 3 is in operation, and vice versa.

Pulse-width modulation is effected by computer control of the gate triggering.

9

A.C. Motor Control.

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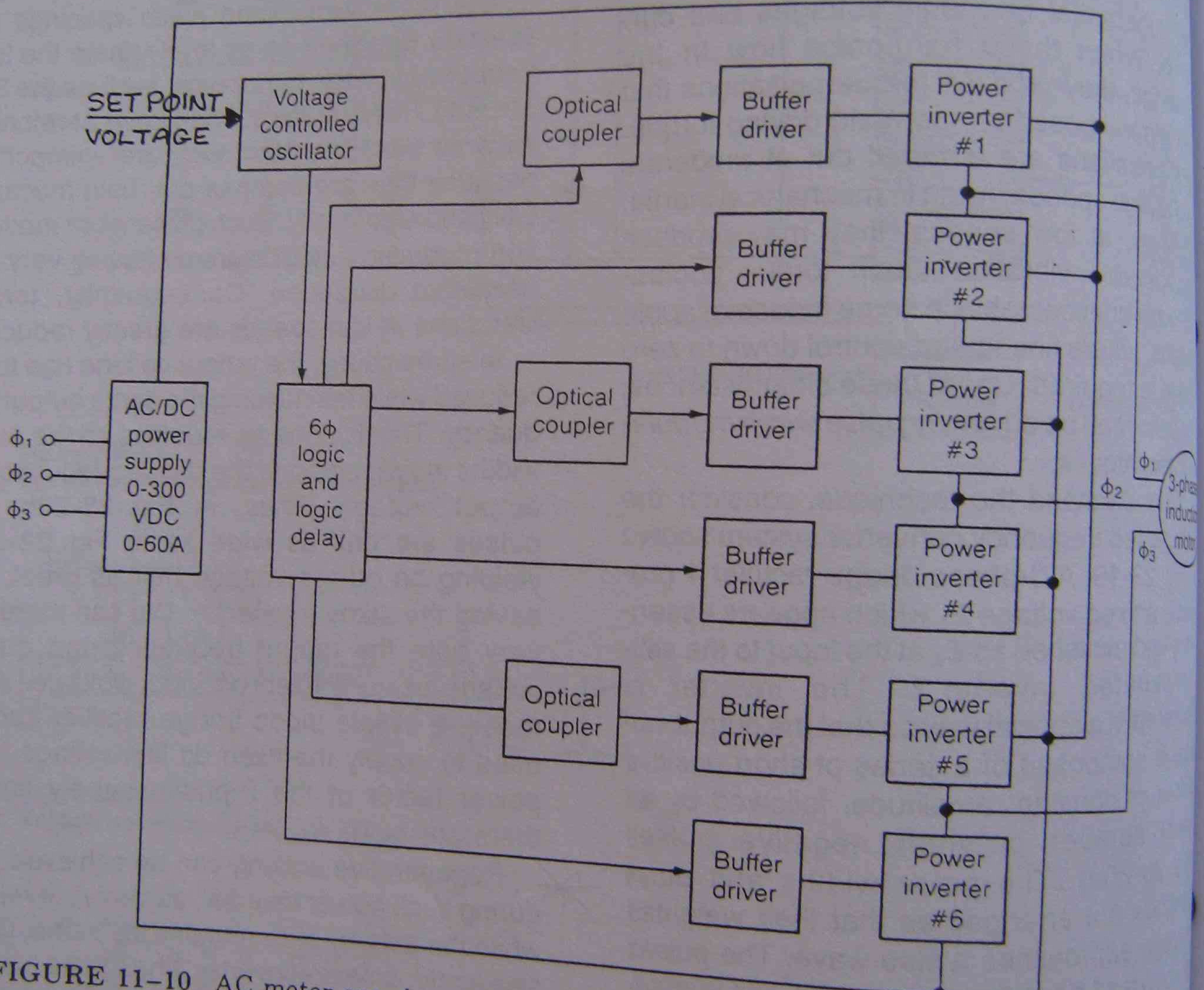

FIGURE 11–10 AC motor speed control—block diagram

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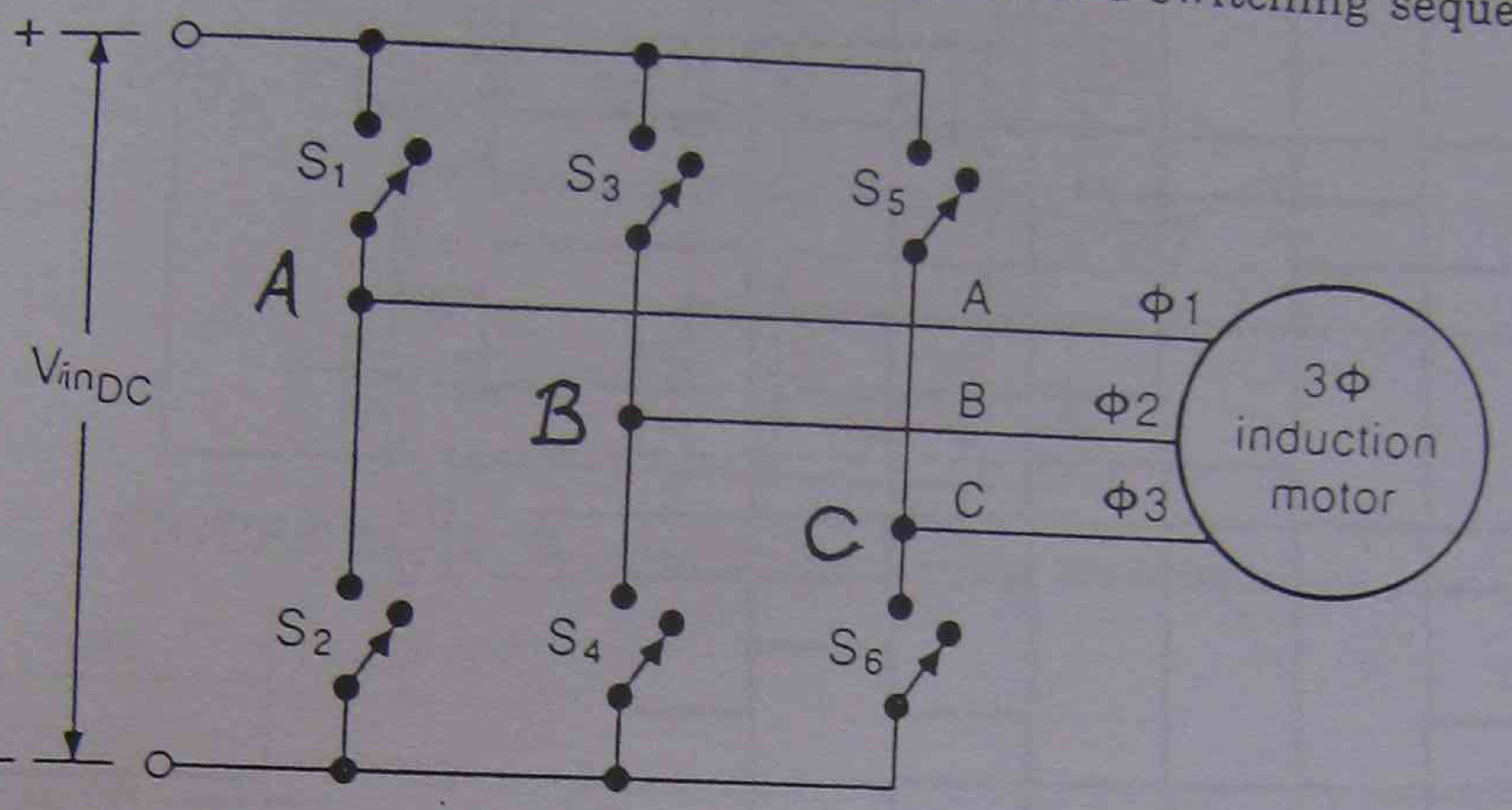
A.C. Motor Control.

Topic -

FIGURE 11–11 Six-step inverter simplified diagram and switching sequence



Period / Switch	1	2	3	4	5	6	7	8
S_1	1	1	1	0	0	0	1	1
S_2	0	0	0	1	1	1	0	0
S_3	0	0	1	1	1	0	0	0
S_4	1	1	0	0	0	1	1	1
S_5	1	0	0	0	1	1	1	0
S_6	0	1	1	1	0	0	0	1

Key – 1 = on 0 = off

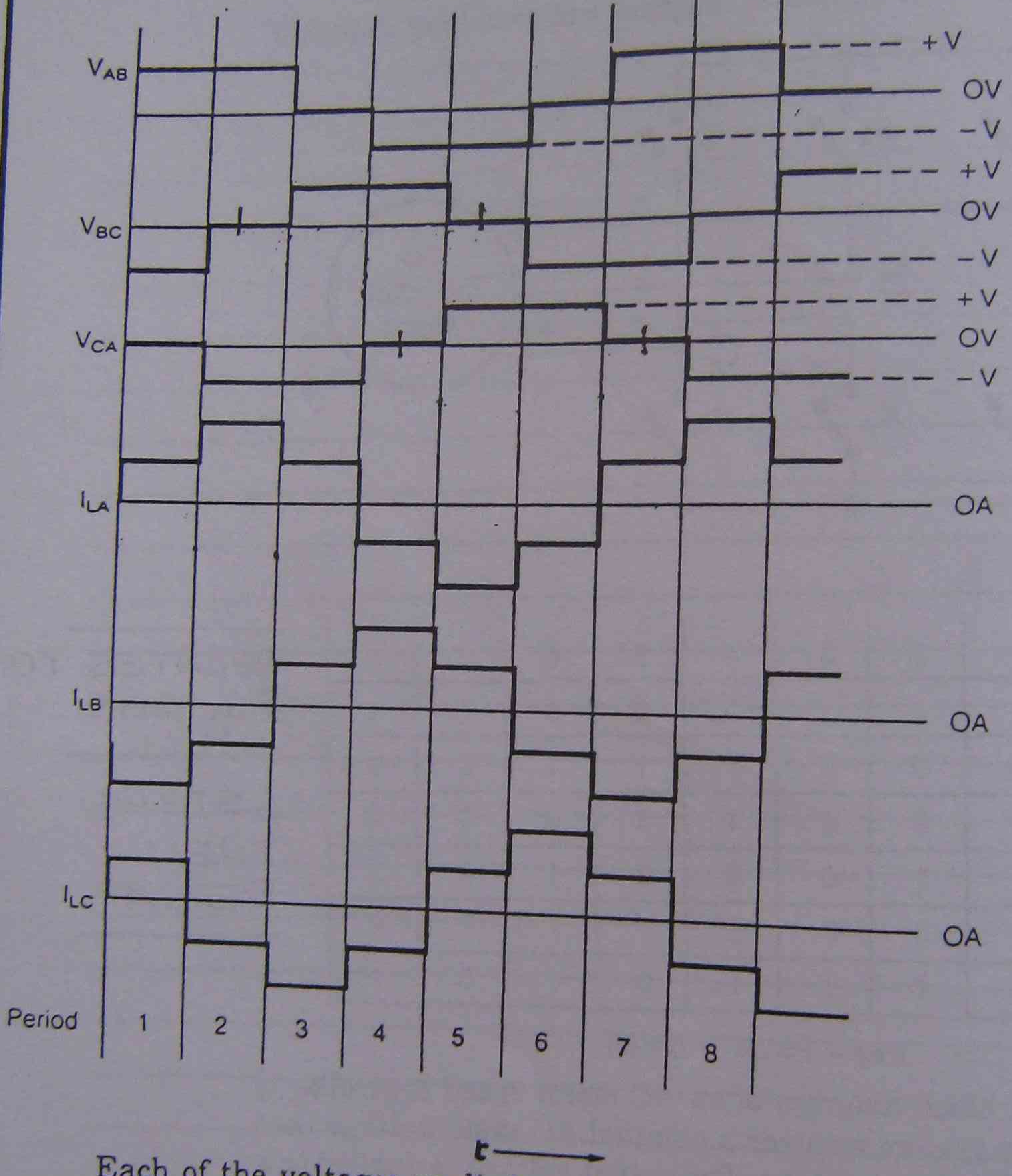
RELATES TO
Fig 11-12

6 STEPS
THEN
REPEAT

Controlling the Motor A block diagram of an AC motor speed controller is shown in Figure 11–10. The system receives a nominal AC input voltage that is converted to a variable DC output voltage. The output voltage is applied to a voltage-controlled oscillator that, in turn, produces a frequency proportional to the DC power supply output voltage. The output of the voltage-controlled oscillator is then used to drive the six-phase logic that will provide properly-timed pulsed outputs to the optical coupler, buffer drivers, and power inverters.

Figure 11–11 shows a simplified six-step inverter diagram that will be used to show the proper switching sequence. Each of the switches shown in Figure 11–11 is actually a transistor or thyristor. The output voltage and current for a resistive load (connected in place of a motor) is shown in Figure 11–12. The current wave forms consist of six distinct steps when the switches are properly sequenced—hence, the name six-step.

FIGURE 11–12 Switching sequence synchrogram



Each of the voltages applied to the three phases is displaced 120° from each other, as shown in Figure 11–12. This figure shows the line-to-line voltages, V_{AB}, V_{BC} and V_{CA}. These voltages were found by adding the voltages algebraically. During periods *1* and *2*, the voltage from *A* to $B = +V$, since *B* is at the $-V$ potential. During period *3*, the voltage *A* to *B* is 0 V, since both *A* and *B* are at +V. In this way, a six-step wave form is achieved.

The output AC voltage can be changed by varying the input DC voltage. The output frequency can be varied by varying the switching frequency of the transistors (*S1* through *S6*). Typically, the maximum frequency for motor speed control using a six-step inverter is 200 Hz.

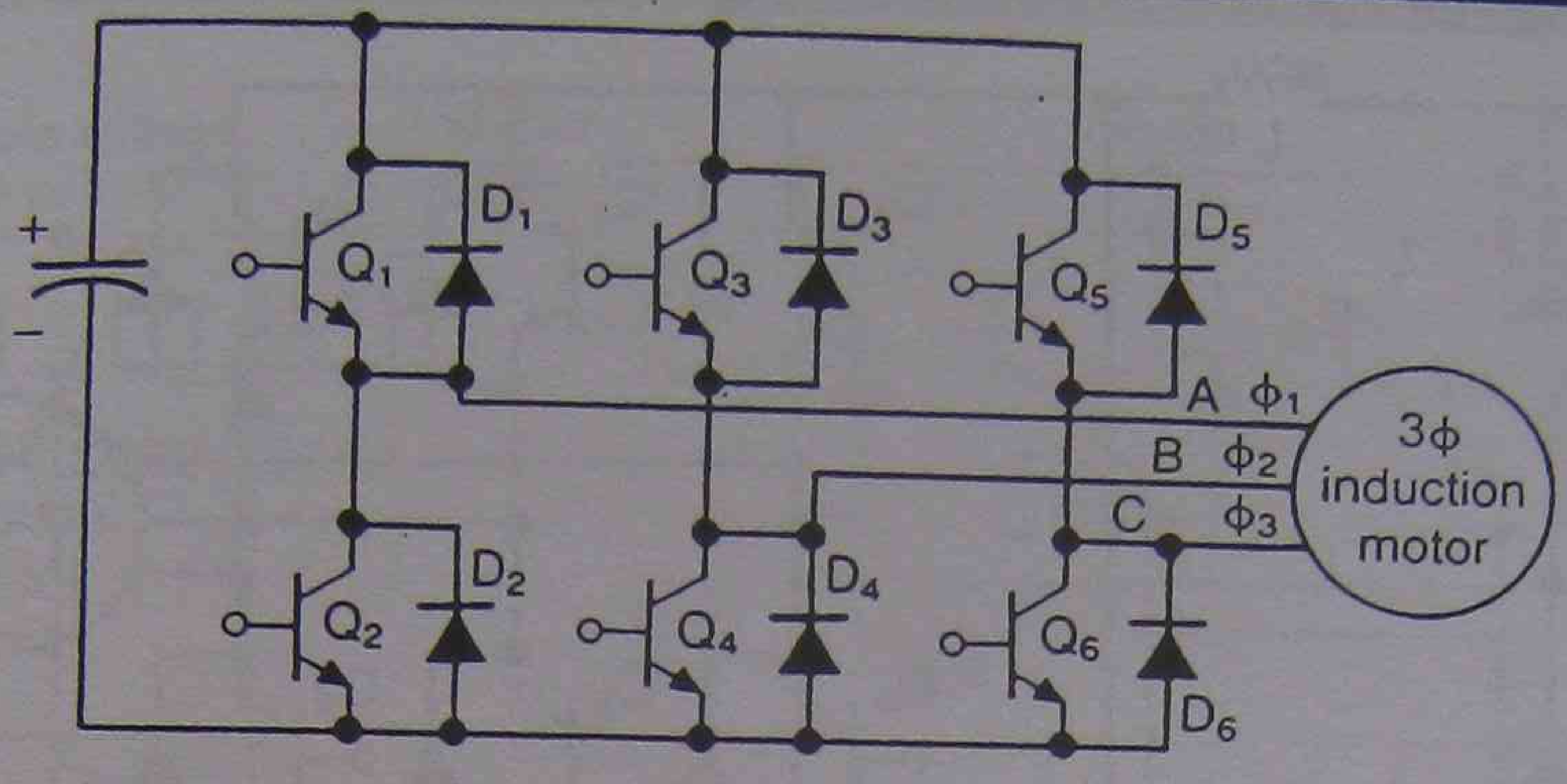


FIGURE 11–13 Six-step inverter using transistors as switches

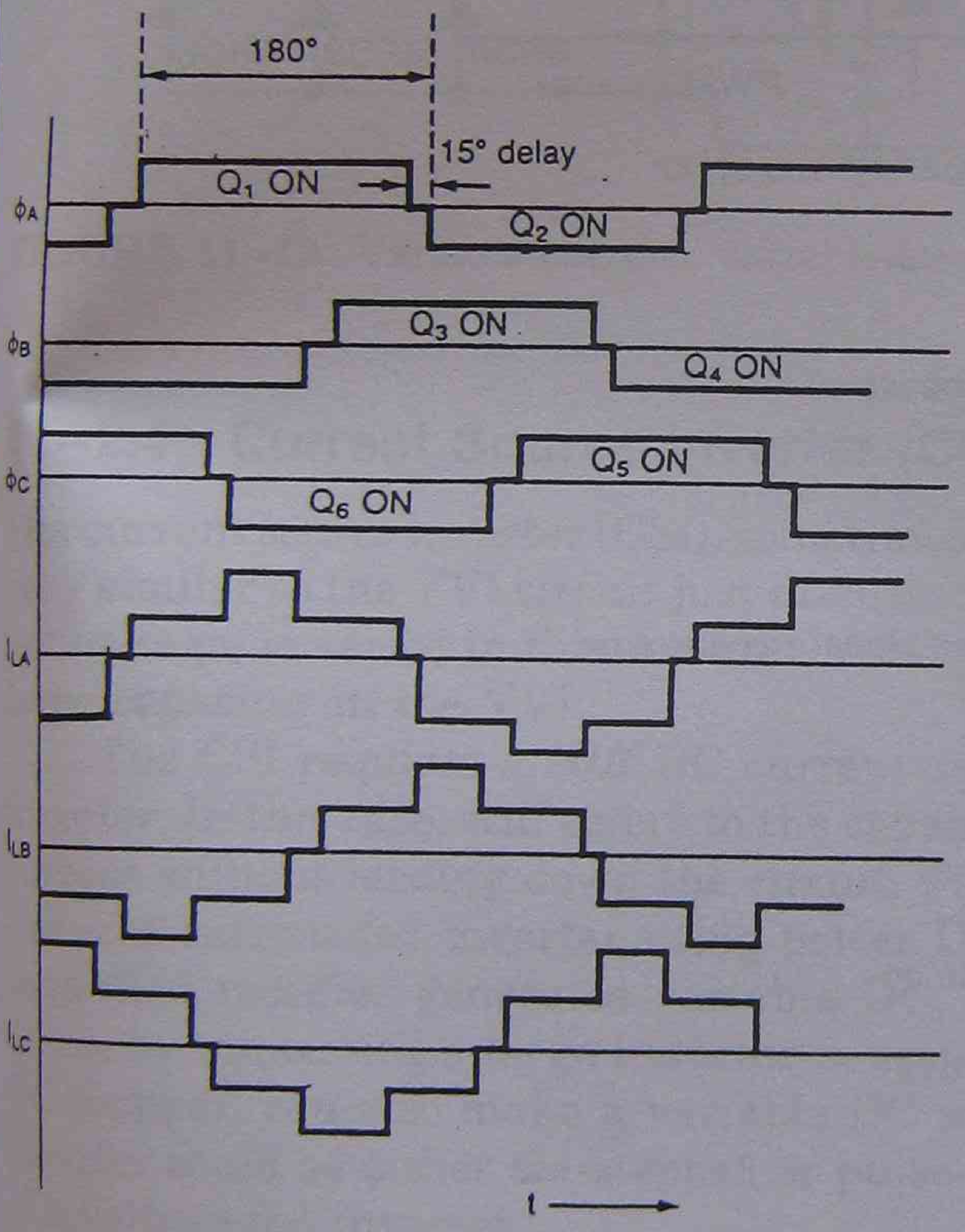


FIGURE 11–15 Typical collector current with motor load

FIGURE 11–14 Six-step inverter synchrogram

Figure 11–12 shows that at any time three switches (transistors) are ducting, one is conducting in each leg of the bridge, and the successive legs switched with delays of 210°. As shown in Figures 11–13 and 11–14, tran- tors *Q1* through *Q6* theoretically conduct for 180°.

However, in a practical situation, it is necessary to provide some time ay (typically 10° to 15°) between the positive-to-negative transition period the phase current. This time delay enables the complementary transistor is complement to Q_2, etc.) to turn off before its opposite member turns on. is action prevents cross-conduction and eventual destruction of the power nsistors. Therefore, the maximum conduction time will be 165° out of a 360° riod. The diodes connected in parallel with each transistor conduct current en the transistor is turned off, represented by −Ic in Figure 11–15.

A.C. Motor Control.

Topic -

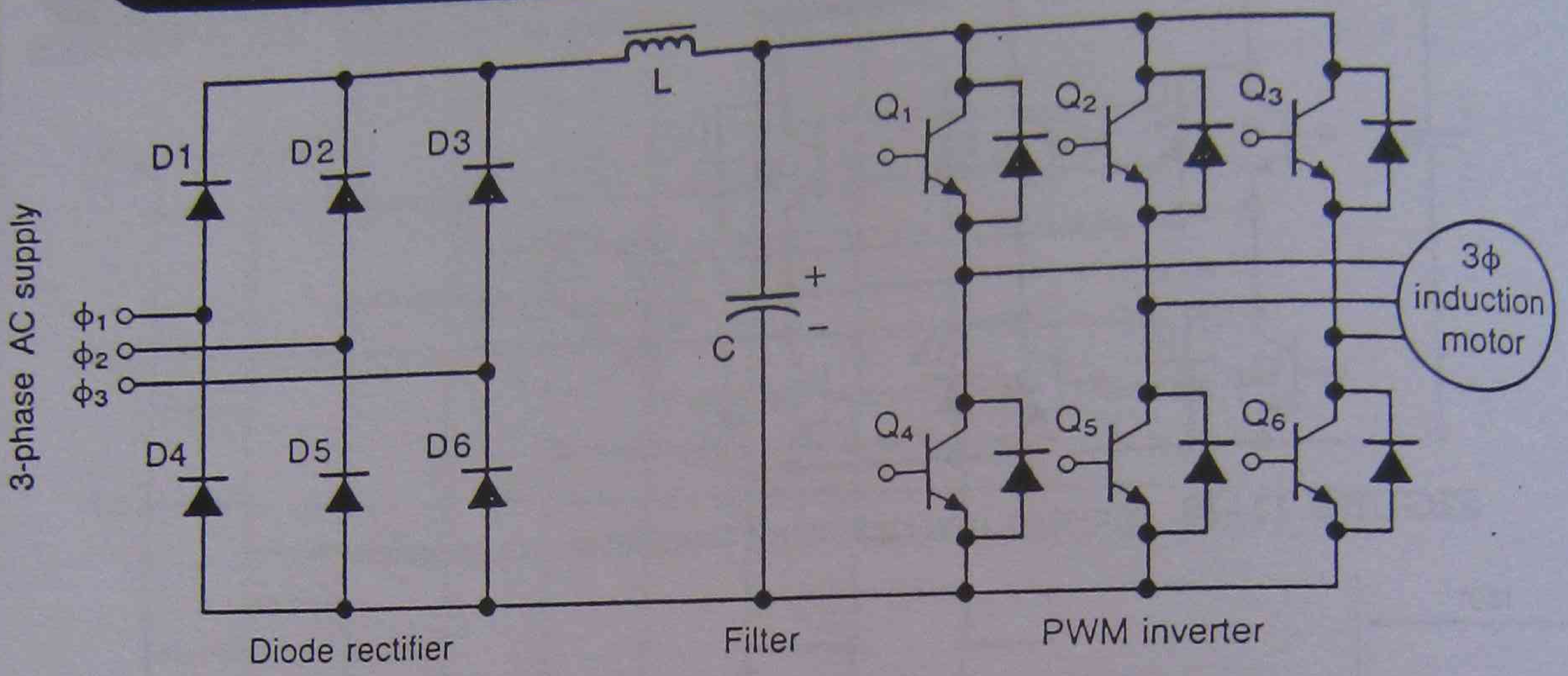


FIGURE 11–17 Variable-voltage, variable-frequency inverter

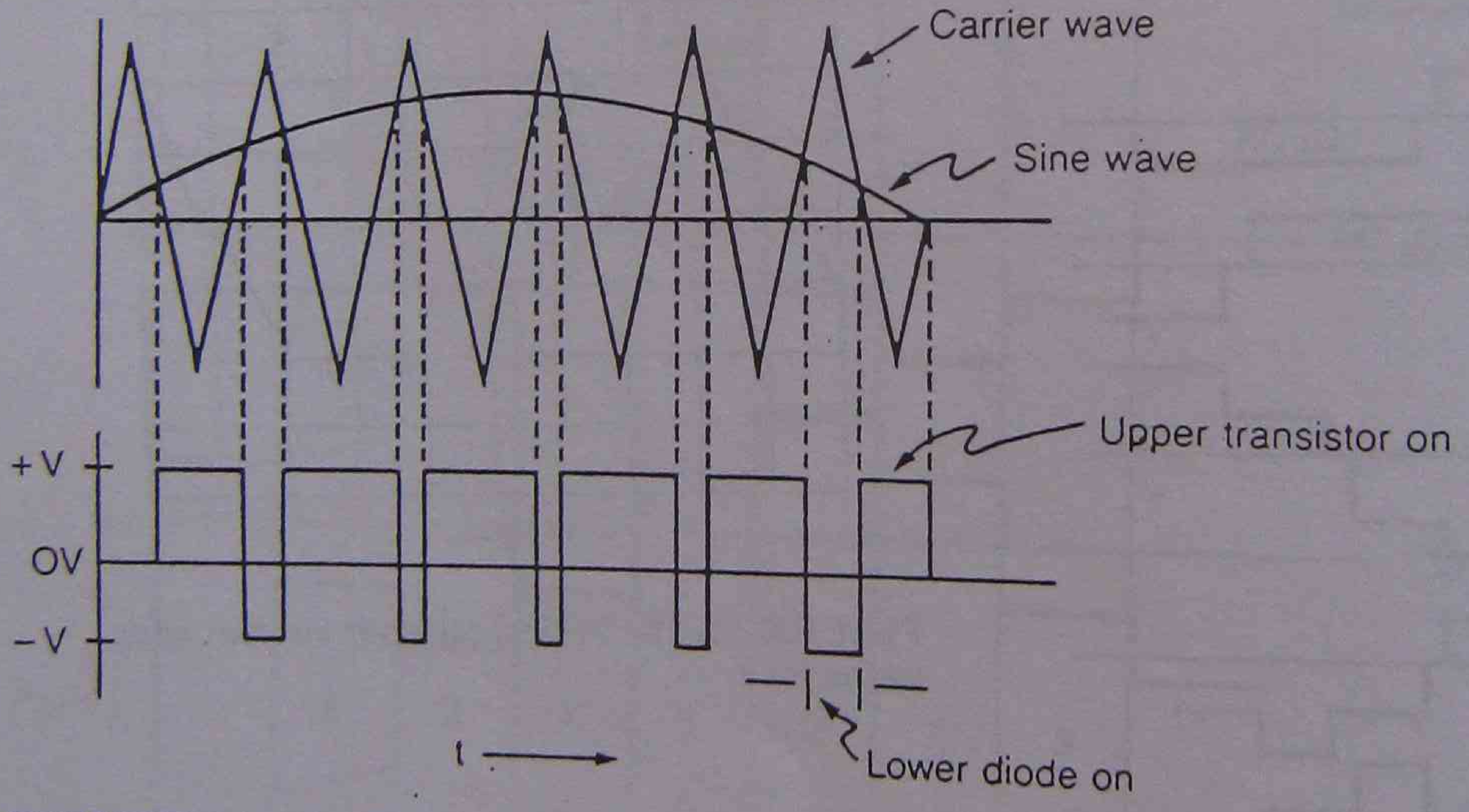


FIGURE 11–18 Pulse-width modulation by sine wave

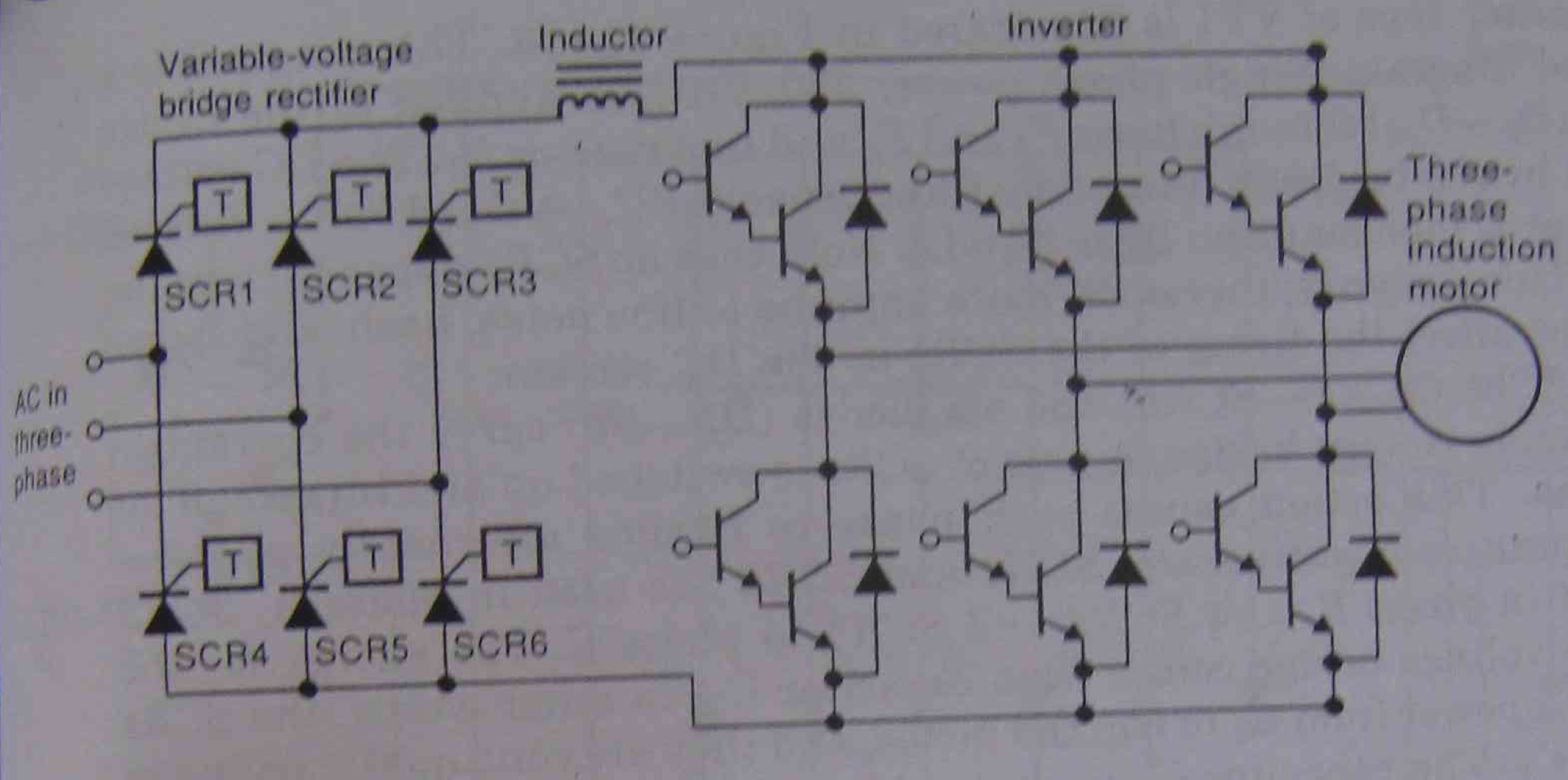


FIGURE 11–19 Variable-current, variable-frequency motor drive

11–2.4 Current Source Inverter (CSI)

The current source inverter (CSI), sometimes called the current-fed inverter, is very similar to the VVI circuit just discussed. Its name suggests that current, not voltage, is varied in this inverter. A large inductor is used in place of the large capacitor in the VVI.

The CSI requires a ***stiff*** DC current source as opposed to a voltage-fed inverter. In this case, stiff refers to the capability to provide a large amount of current without loading down the circuit. Figure 11–19 shows the power circuit of a current-fed inverter using power Darlingtons as switches. A phase-controlled rectifier generates variable DC, which is converted to a current source by connecting a large inductor in series. A diode rectifier, followed by a DC chopper, can also make a variable DC source. The mode of control of the inverter could be either six-stepped or pulse-width modulated, similar to that of a voltage-fed inverter.

Another type of VVI is illustrated in Figure 11–16. This diagram is a simplified diagram. Single-phase power, 220 VAC, is applied to the bridge rectifier (D_9—D_{12}) through fuses F_1 and F_2 and line reactor L_1. The DC voltage produced by the bridge is filtered by filter capacitor C_4. A fixed value, filtered DC voltage is then found on lines *1* and *5*. Note that no SCRs appear across the AC line. This design is, therefore, more immune to line noise, hash, and spikes that might affect the firing of the SCRs in the DC section.

Six SCRs (*SCR1—SCR6*) and six diodes (*D1—D6*) carry the current in the adjustable voltage bridge. A pair of SCRs is switched on and turned off for each phase. This action causes each phase to become alternately positive-negative-positive-negative, etc. *SCR1* and *SCR2* are used in phase *A*, *SCR3* and *SCR4* in phase *B*, and *SCR4* and *SCR5* in phase *C*. The energy for the adjustable voltage bridge comes from capacitor C_3. In other words, the SCRs are drawing power from C_3 to run the motor, and they are continually trying to discharge C_3. The capacitor C_3 is charged by the fixed voltage bridge.

Each pair of SCRs controlling each phase is turned on and commutated off in the proper timing sequence to supply three-phase power to the motor by the main control circuit board (not shown in the simplified diagram). The faster the switching on and commutating off, the higher the frequency.

The six SCRs in the fixed voltage bridge (*SCR11—SCR16*) operate in parallel with their equivalents in the adjustable voltage bridge (*SCR1—SCR6*). The purpose of the fixed voltage bridge is to furnish energy to C_3 as the motor load uses it and to commute (turn off) the SCRs in the adjustable voltage bridge.

When an SCR in the adjustable voltage bridge, for example *SCR1*, is turned on, its equivalent in the fixed voltage bridge (*SCR11*) can be turned on at 10 kHz rate. It will be turned on for one pulse any time C_3 voltage is too low. When it is turned on for a pulse, energy comes from the DC line (lines *1* and *5*) through X_1 or X_2. Some of this energy goes to the motor, and the excess goes through the back diode (*D1—D6*) and helps recharge C_3. If C_3 voltage continues to be too low, another pulse of energy will be called for and once again the SCR in the fixed voltage bridge will be turned on. Up to 10,000 pulses of energy per second can be obtained from each SCR in the fixed voltage bridge in this manner.

To commutate off an SCR in the adjustable voltage bridge, for example *SCR1*, its gate is turned off. Then its mate in the fixed voltage bridge (*SCR11*) is turned on for a pulse, causing current to flow. Some current flows to the motor; the excess flows through the back diode *D1*. This action shunts *SCR1*, causing it to stop conducting.

The SCRs in the fixed voltage bridge turn themselves off naturally after each pulse because they conduct into a tuned circuit. The voltage rises at the end of the pulse to block any further current flow.

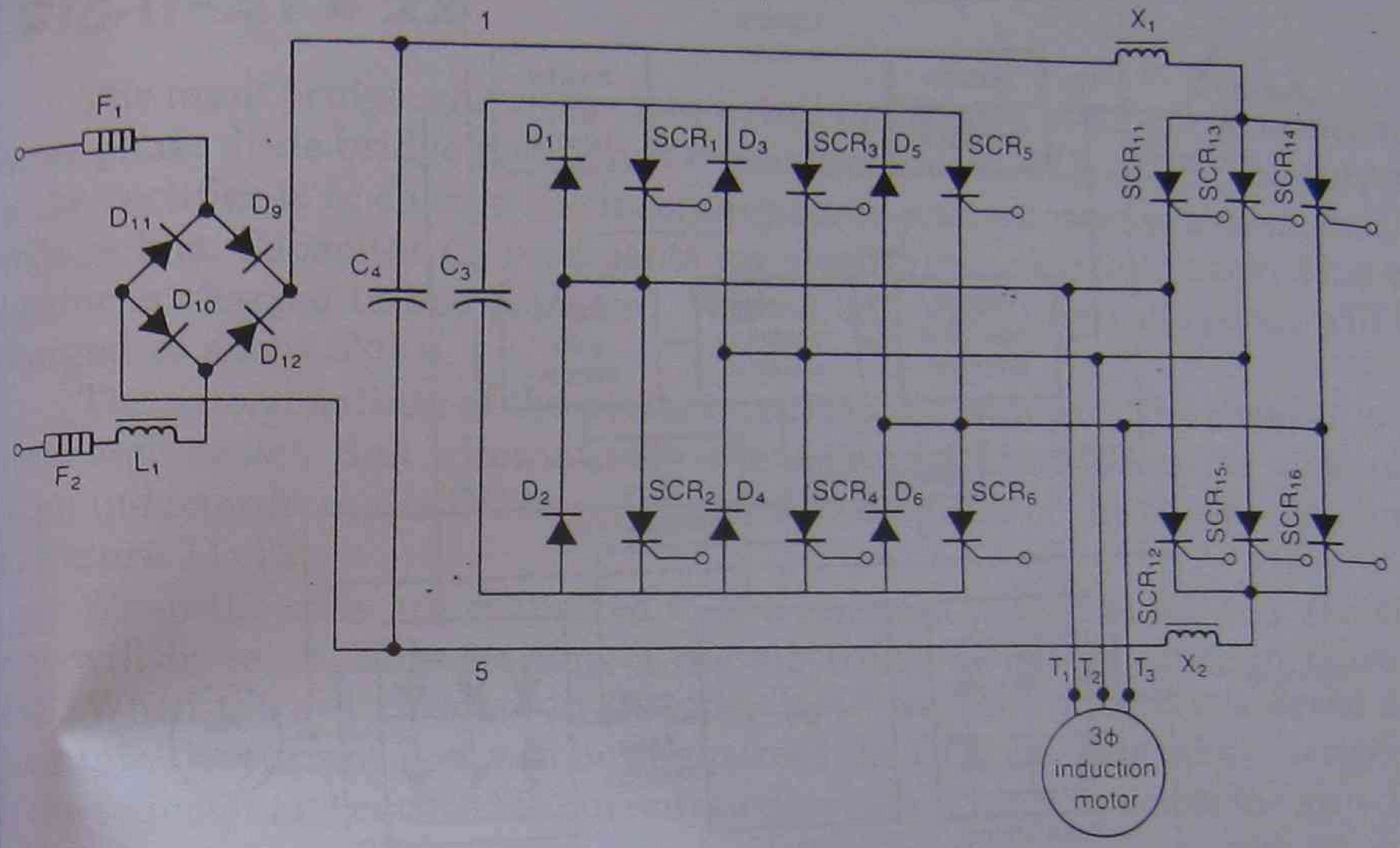
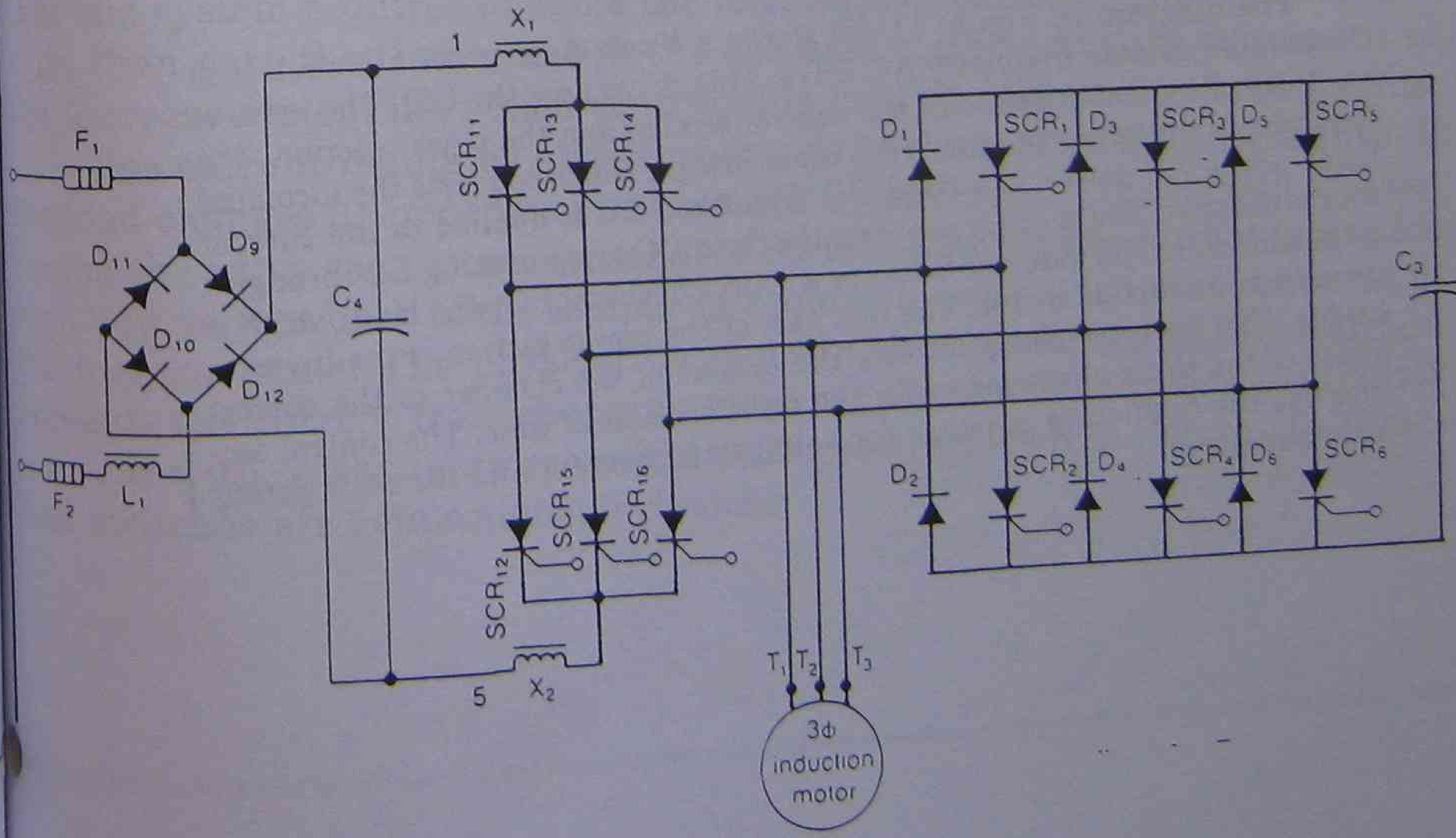


FIGURE 11–16 VVI AC motor drive using SCRs as switches



A.C. Motor Control.

Topic -

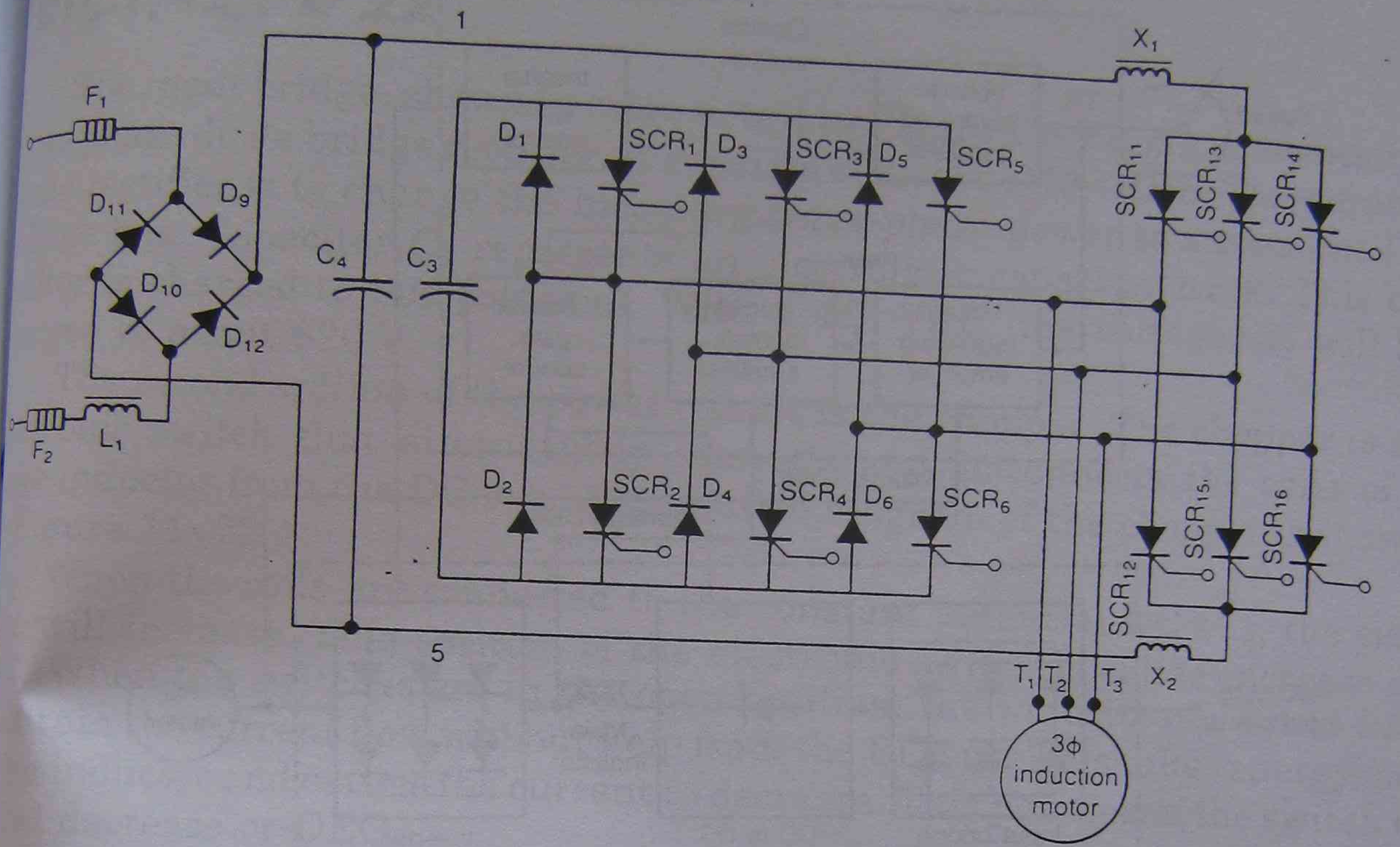
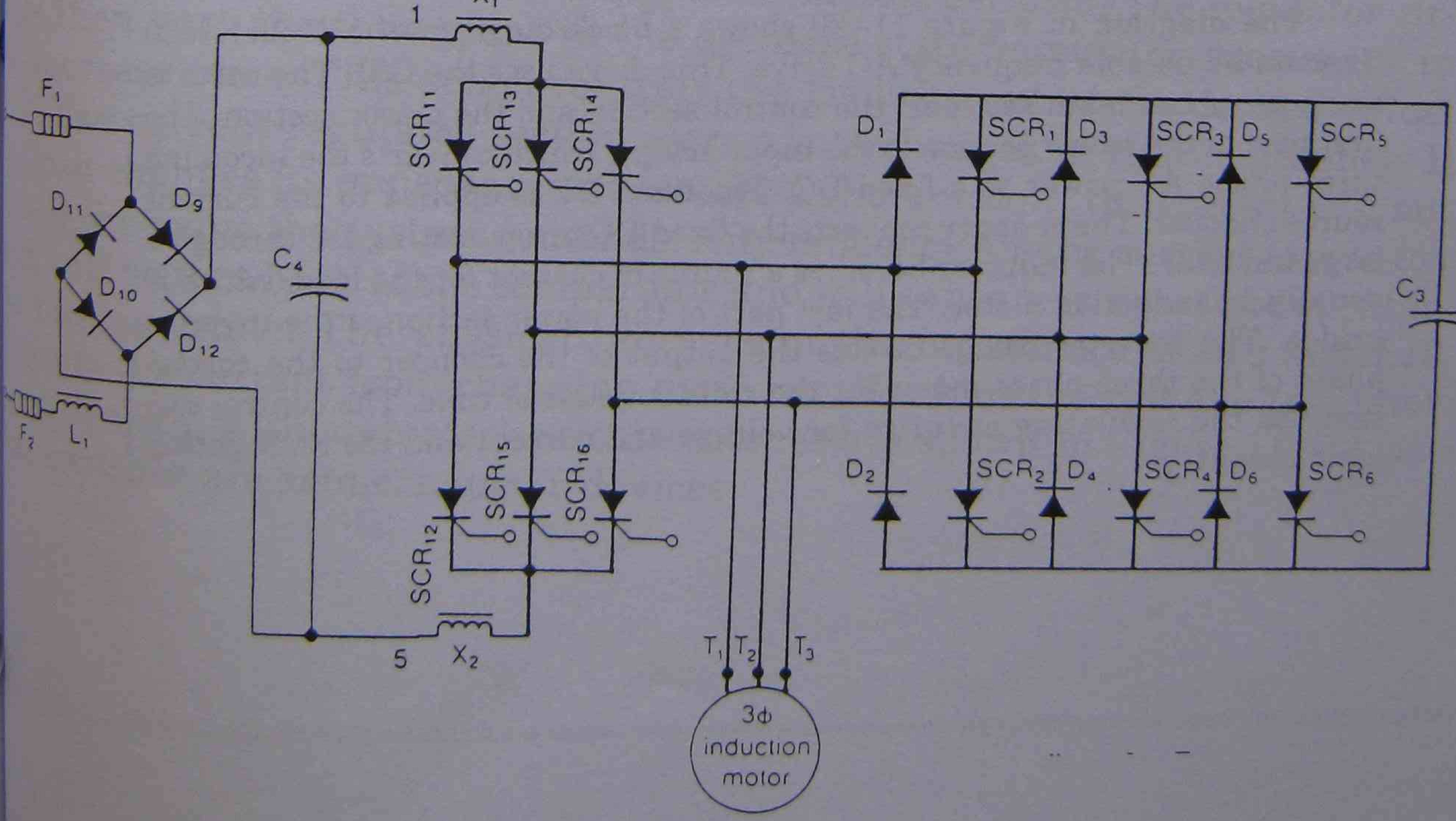


FIGURE 11–16 VVI AC motor drive using SCRs as switches



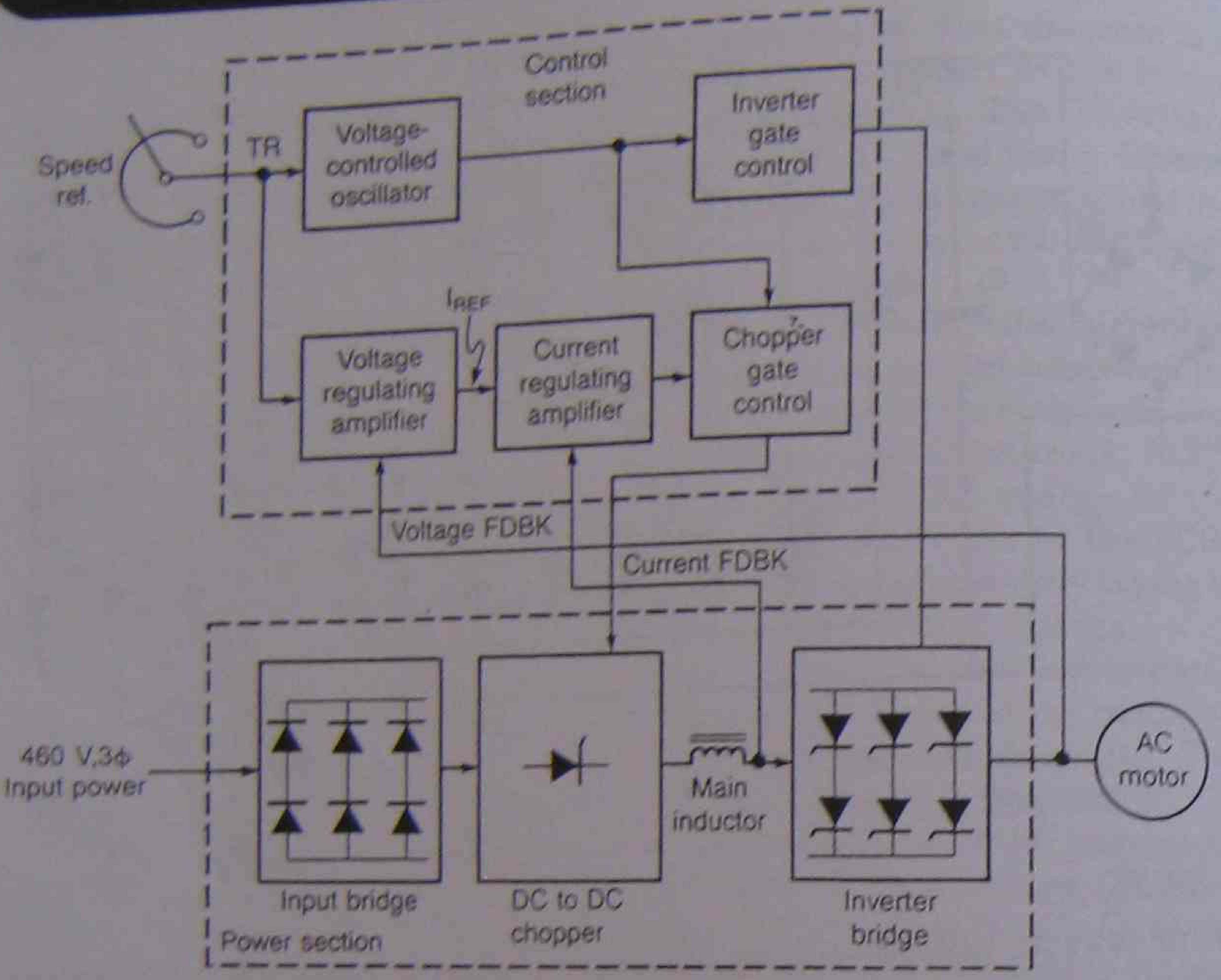


FIGURE 11–20 Graham 1580 CSI AC motor speed control (courtesy of Graham Co.)

The diagram in Figure 11–20 shows a block diagram of a Model 1580 Graham adjustable frequency AC drive. This drive uses the CSI. The drive is made up of two main sections: the control section and the power section. The first part of the power section is the input bridge, which converts the incoming three-phase AC power to a fixed DC. The fixed DC is applied to the current source chopper. The chopper converts the fixed DC to a pulsating DC through a large inductor. The inductor becomes a source of current for the load, which is usually an induction motor. The last part of the power section is the inverter bridge. The inverter bridge directs the output of the chopper to the correct phase of the three-phase motor for the right amount of time. The control section has the regulating circuitry for voltage and current and the SCR gating logic circuitry.

FIG 11-21 & 22

The input bridge, shown in more detail in Figure 11–21, is a conventional three-phase diode bridge rectifier. It consists of diodes *D13—D18*. The purpose of the rectifier is to change the incoming three-phase power to a constant DC voltage bus. Capacitor C_1 represents an electrolytic capacitor bank. This capacitor is charged to bus potential. With a 460 VAC line voltage, C_1 will be charged to about 620 V.

The second section of the power circuit is the chopper. The chopper is an electronic switch that alternatively connects and disconnects the coils of a large inductor from the DC bus. A simplified diagram of the chopper is shown in Figure 11–22.

When the coils are connected to the constant potential DC bus, the current will increase. This position of the electronic switch is called increase or INC. When the switches are in the open position, the inductors *L18* and *L19* maintain the current flow without help from the DC bus. This takes energy out of the inductor and causes the current to decrease. This position of the switch is called decrease or DEC.

The load current is sensed by a Hall-effect device, which gives an output voltage proportional to the flux created by the load current. This proportional signal is then fed back to the regulator circuit where it is compared to a value that the system requires to make the load (motor) perform. The regulator circuits then vary the time spent in the decrease state versus the time spent in the increase state. This action controls the load current at the desired value.

The current-regulated section responds to changes in load impedance. If the load changes its impedance or becomes a short circuit, the current will not exceed the value the system demands. Any tendency for the current to rise too high will be sent back to the comparator and will cause less time to be spent in the increase state. The physical presence of the inductance of *L18* and *L19* prevents the current from changing more rapidly than the rate the regulator can cope with. Although mechanical switches are shown in Figure 11–22, actual switches are semiconductor devices.

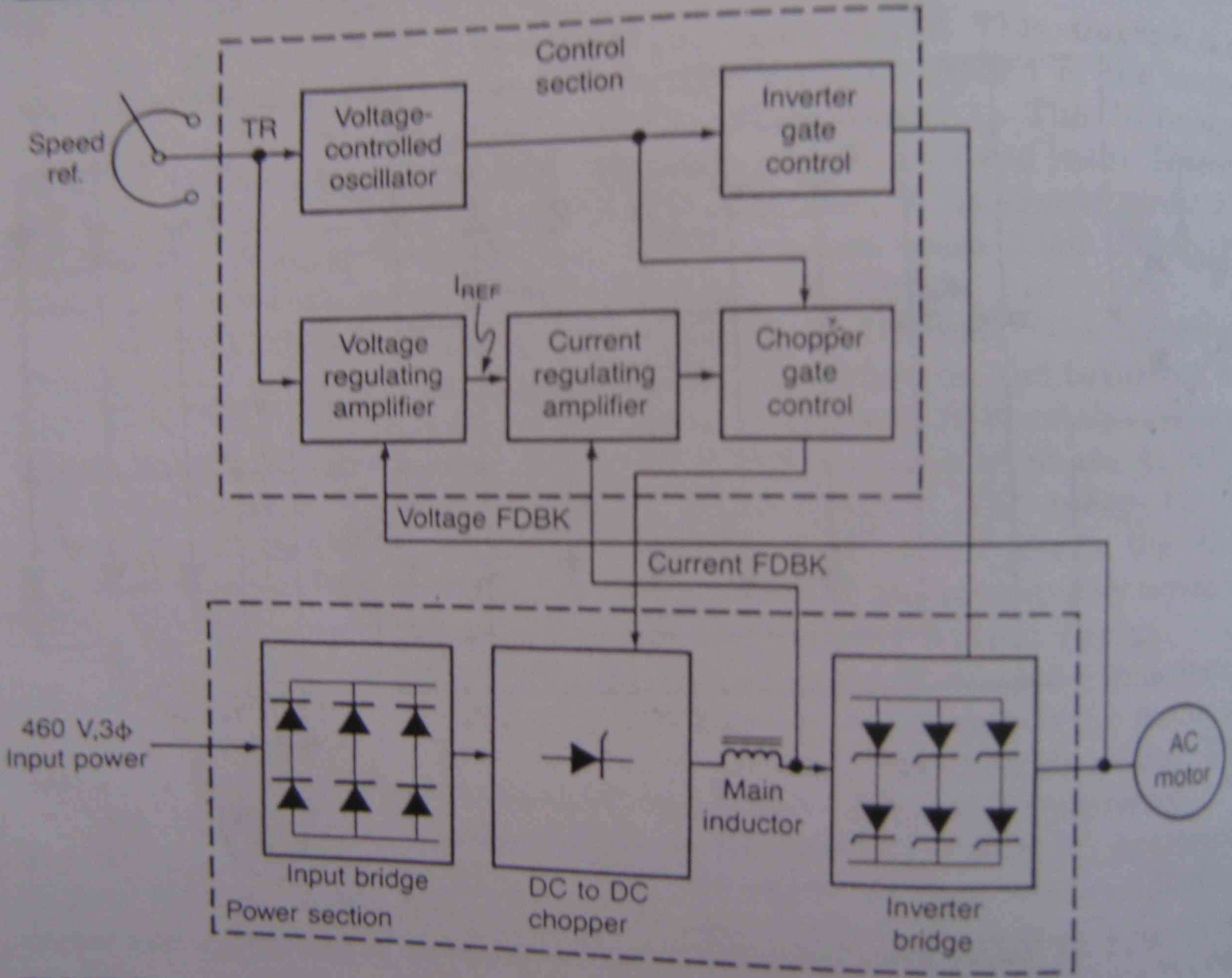


FIGURE 11–20 Graham 1580

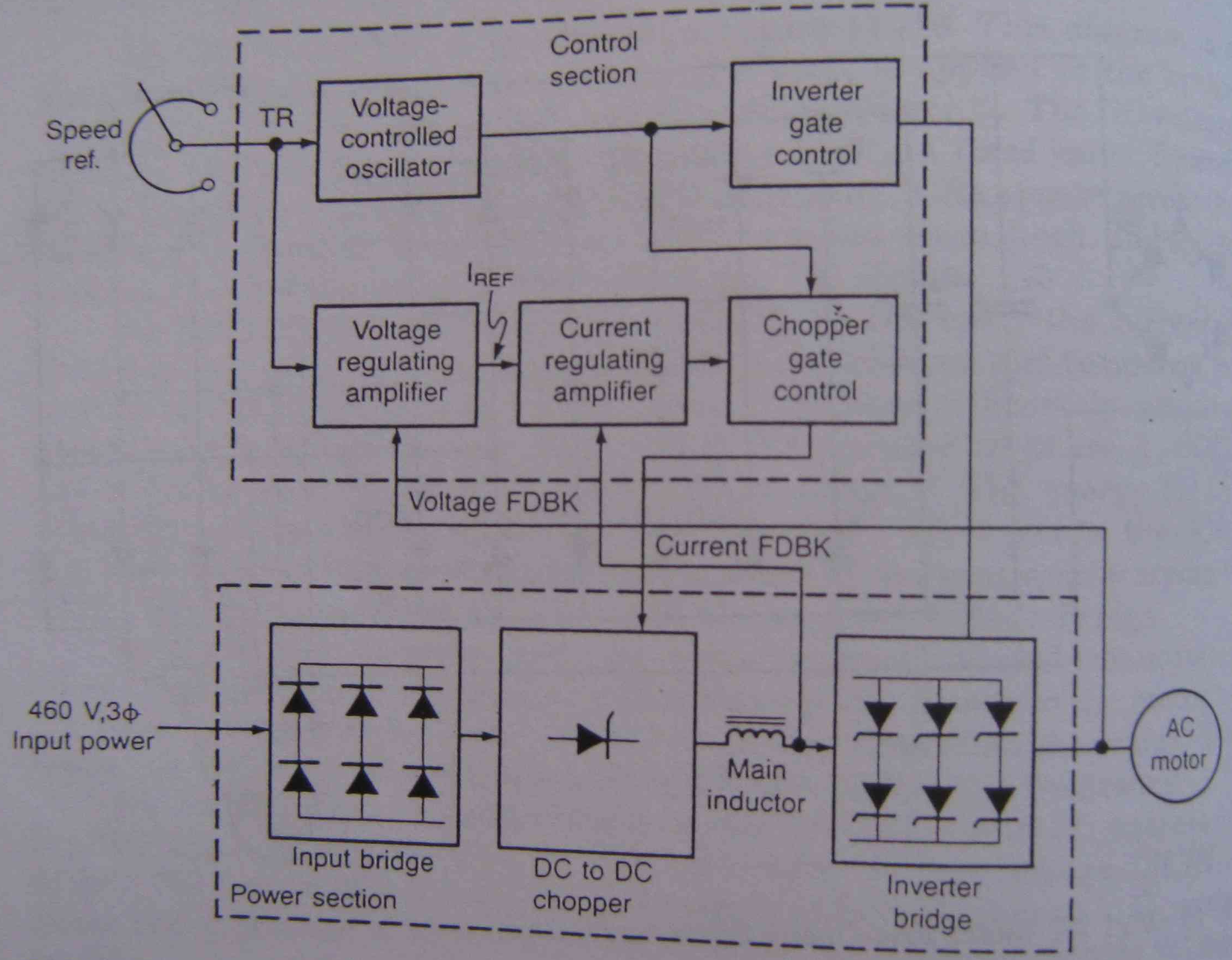


FIGURE 11–20 Graham 1580 CSI AC motor speed control (courtesy of Graham Co.)

The diagram in Figure 11–20 shows a

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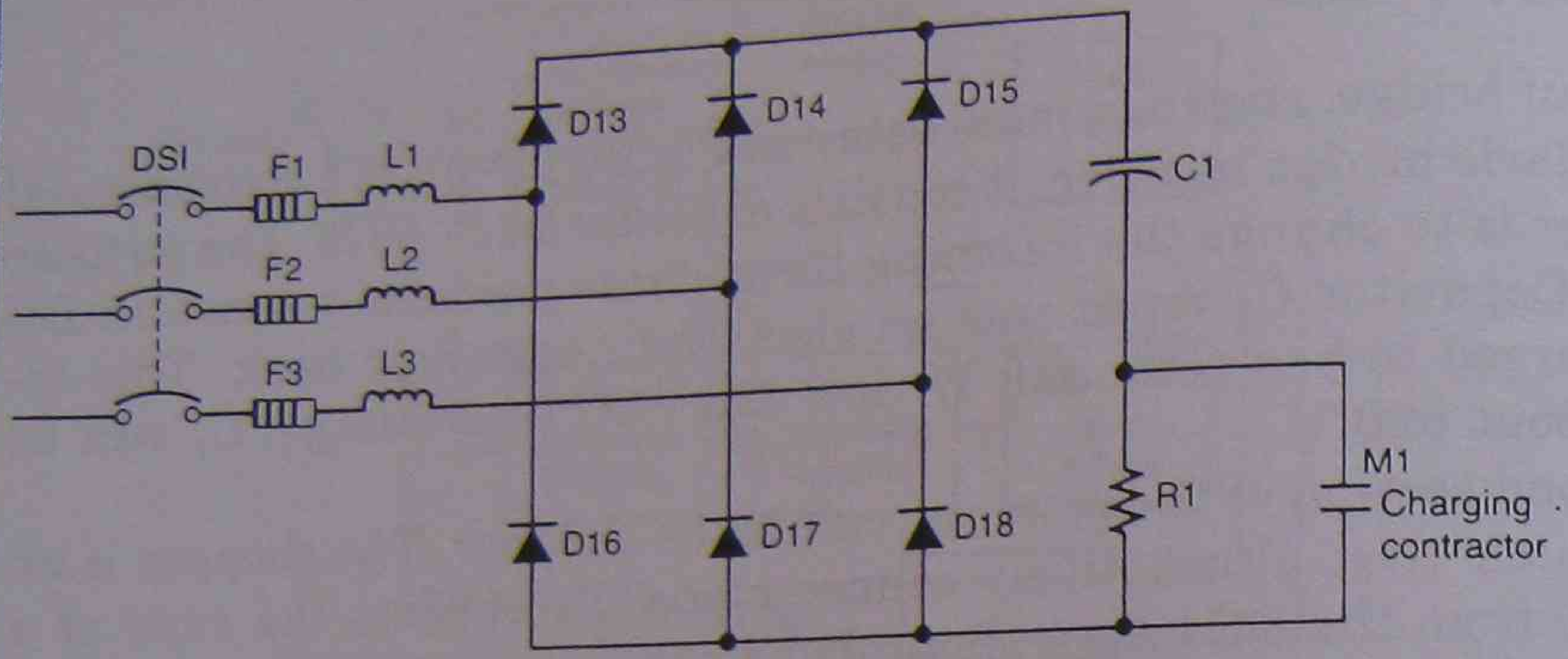


FIGURE 11–21 Input bridge, Graham CSI control (courtesy of Graham Co.)

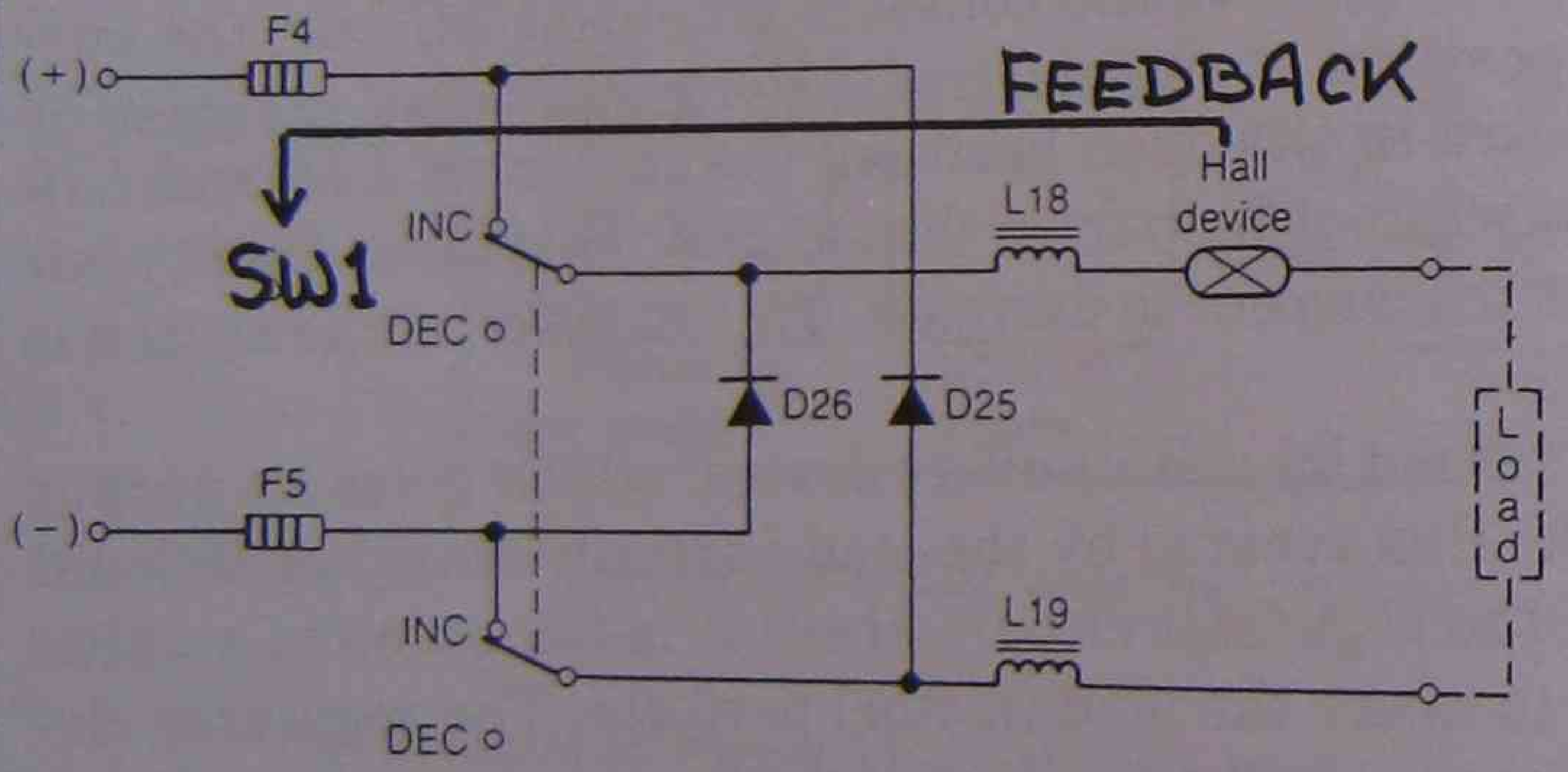


FIGURE 11–22 Chopper circuit, Graham CSI controller (courtesy of Graham Co.)

FIG 11-23

The schematic in Figure 11–23 is a simplified schematic of the inverter section. It is a six-step inverter with *SCR1–6* switching the load current at the proper rate as determined by the control circuitry. The switching rate of the SCRs establishes the output frequency. Commutation capacitors *C28* through *C33* store energy necessary for turning off the SCRs by reversing their terminal voltage. Series diodes *D1* through *D6* isolate the capacitors from the load. Only two SCRs are on at any one time, with each one conducting the 120°. They are commutated when the adjacent SCRs in the next phase are fired i the order numbered.

The inductance of the motor is a significant factor in the commutating scheme. The inductance stores energy necessary for commutation, and thi energy charges the capacitors for the next cycle. There is also a precharg circuit consisting of a diode and resistance in series with each capacitor. Th precharge circuit is present to insure that enough capacitor voltage is presen to commutate the SCRs during starting and low frequency operation.

The chopper controls the current in the inverter coming from *L18*. Eve when there are two parallel paths for current, the sum of the two can never b greater than the output current from the chopper. The inverter controls onl the time that the current flows through each motor phase.

The DC voltage at the input terminals of the inverter will vary with th demand of the load. At no load, the voltage will be near zero. At rated load, will be maximum.

The response of the motor and the load and the applied current and fre quency determine the counter EMF (CEMF) of the motor. It is approximatel sinusoidal with a spike occurring each time an SCR is commutated. Figure 11 24 shows a typical phase current and voltage in the inverter section.

We can see that current flows for 120° in the positive direction, ceases f 60° and repeats, but in the negative direction. The positive current for phase is when *SCR1* conducts with either *SCR6* or *SCR2*. Negative current is draw when *SCR4* conducts with either *SCR3* or *SCR5*. The sawtooth on top of th current coincides with the chopper changing between the increase and d crease states.

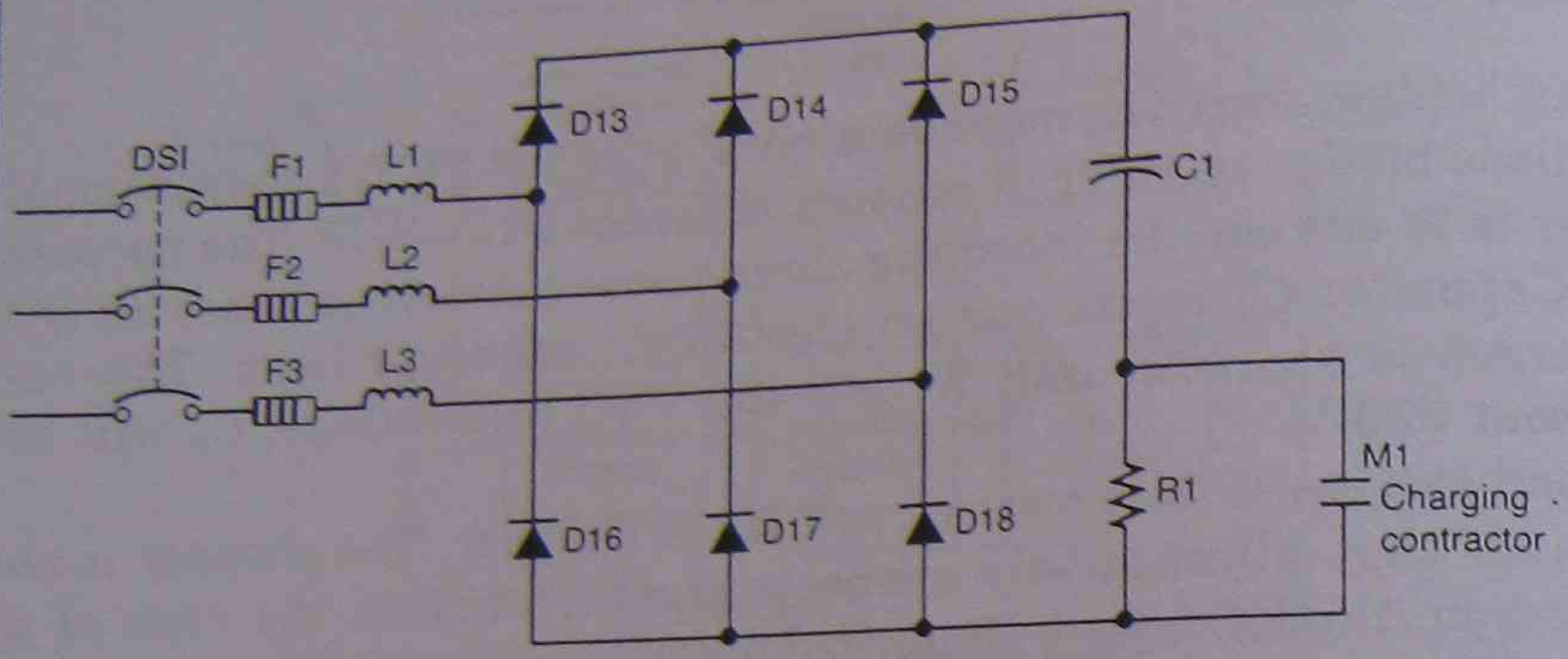


FIGURE 11–21 Input bridge, Graham CSI control (courtesy of Graham Co.)

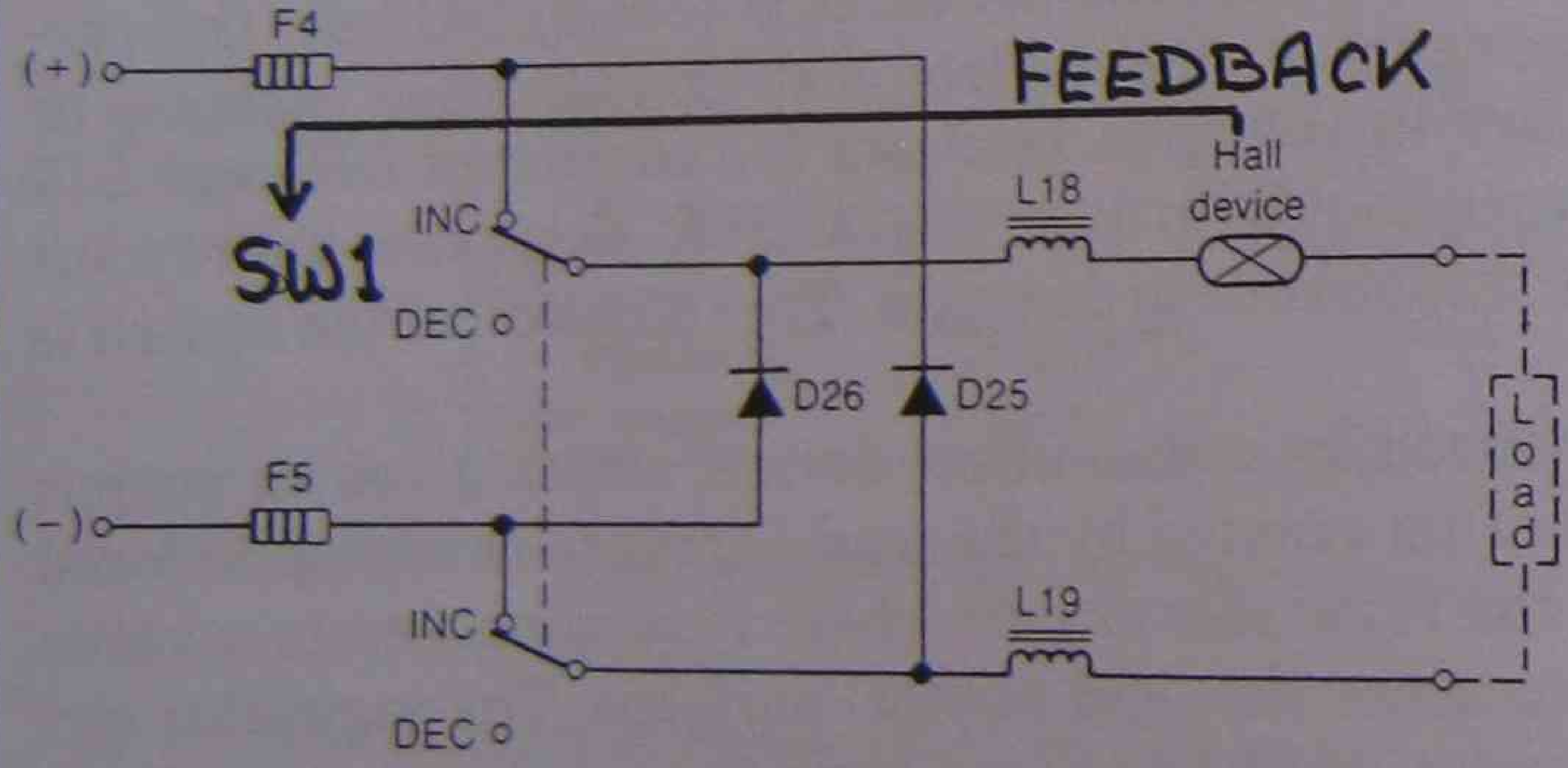


FIGURE 11–22 Chopper circuit, Graham CSI controller (courtesy of Graham Co.)

FIG 11-23

The schematic in Figure 11–23 is a simplified schematic of the inverter section. It is a six-step inverter with *SCR1–6* switching the load current at the proper rate as determined by the control circuitry. The switching rate of the SCRs establishes the output frequency. Commutation capacitors *C28* through *C33* store energy necessary for turning off the SCRs by reversing their terminal voltage. Series diodes *D1* through *D6* isolate the capacitors from the load. Only two SCRs are on at any one time, with each one conducting the 120°. They are commutated when the adjacent SCRs in the next phase are fired in the order numbered.

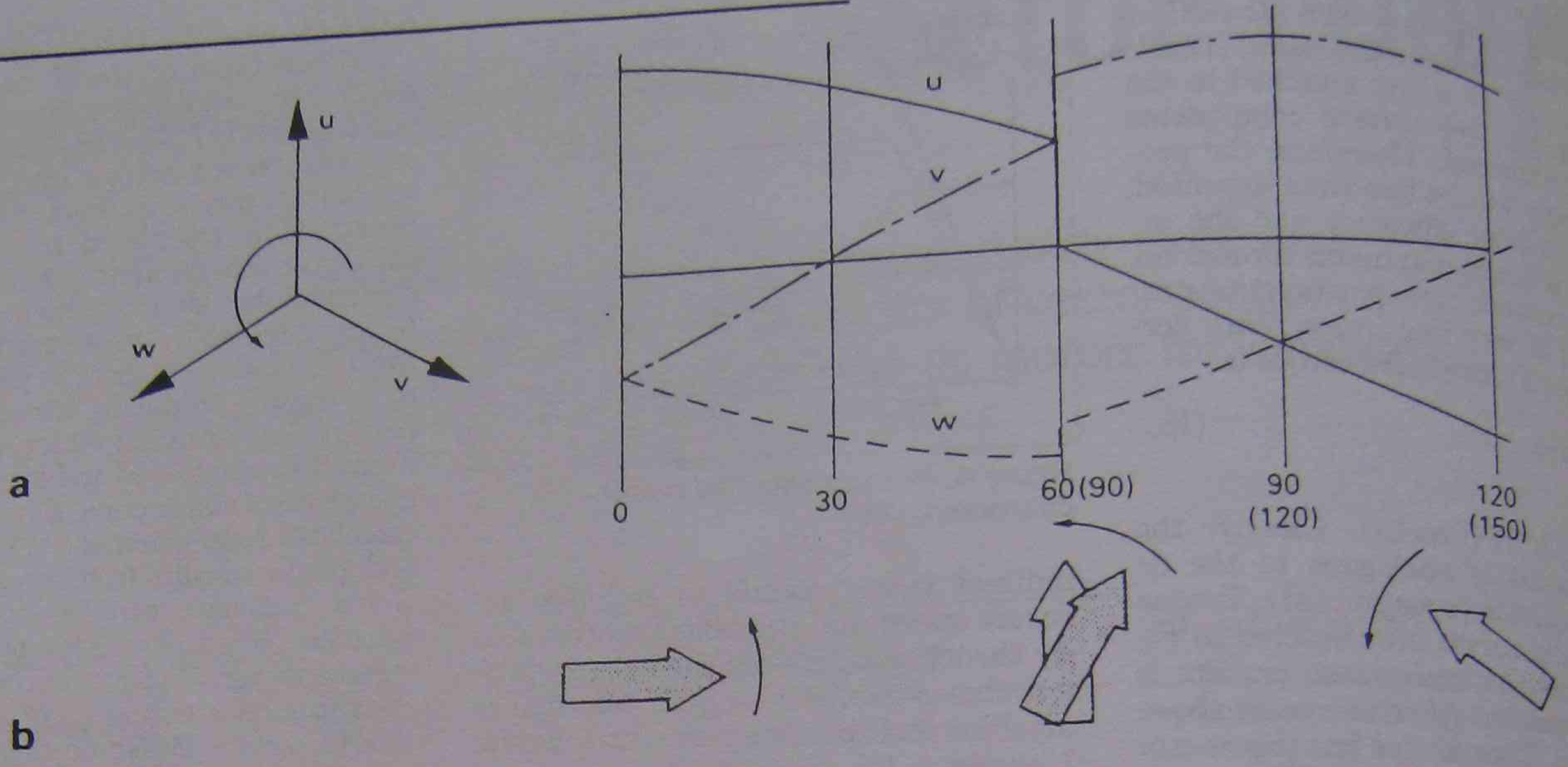
The inductance of the motor is a significant factor in the commutating scheme. The inductance stores energy necessary for commutation, and this energy charges the capacitors for the next cycle. There is also a precharge circuit consisting of a diode and resistance in series with each capacitor. The precharge circuit is present to insure that enough capacitor voltage is present to commutate the SCRs during starting and low frequency operation.

The chopper controls the current in the inverter coming from *L18*. Even when there are two parallel paths for current, the sum of the two can never be greater than the output current from the chopper. The inverter controls only the time that the current flows through each motor phase.

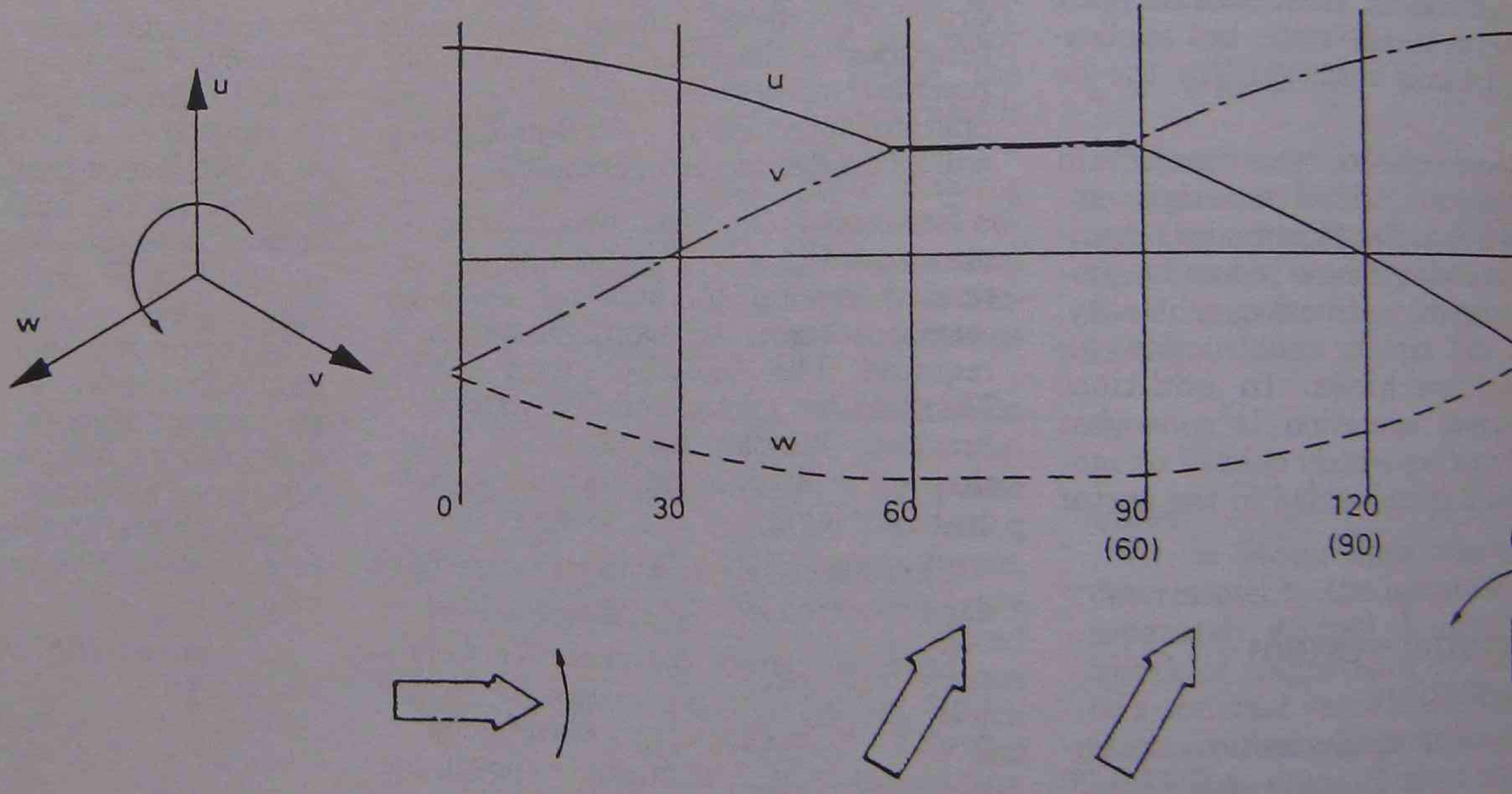
The DC voltage at the input terminals of the inverter will vary with the demand of the load. At no load, the voltage will be near zero. At rated load, it will be maximum.

The response of the motor and the load and the applied current and frequency determine the counter EMF (CEMF) of the motor. It is approximately sinusoidal with a spike occurring each time an SCR is commutated. Figure 11–24 shows a typical phase current and voltage in the inverter section.

We can see that current flows for 120° in the positive direction, ceases for 60° and repeats, but in the negative direction. The positive current for phase *A* is when *SCR1* conducts with either *SCR6* or *SCR2*. Negative current is drawn when *SCR4* conducts with either *SCR3* or *SCR5*. The sawtooth on top of the current coincides with the chopper changing between the increase and decrease states.



2 *Step-change of field rotation:* **a** *Phase waveform;* **b** *Reorientation of flux vectors.*



3 *Stationary field flux vector.*

TABLE 1
3-Phase Rotating Field — Principal Variations

3-phase Current Vectors	Stator Flux Vector
Normal Mains Supply	
Balanced, sinusoidal, constant frequency and amplitude	Constant magnitude, rotating at constant angular velocity
PWM Supply Changes	
Character of Change:	**Effect:**
Amplitude	Magnitude of current vectors
Phase sequence	Directio

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A.C. Motor Control

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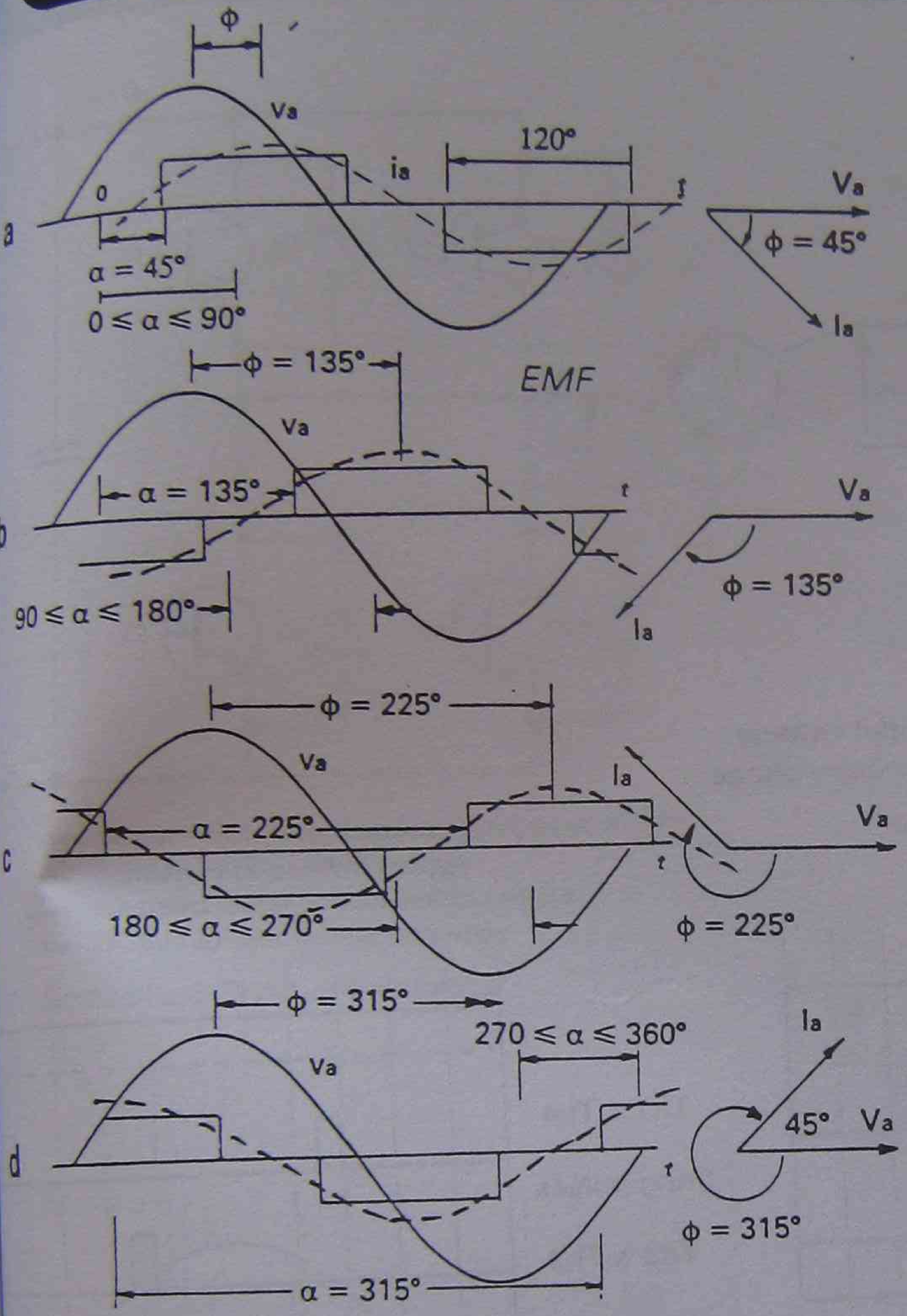
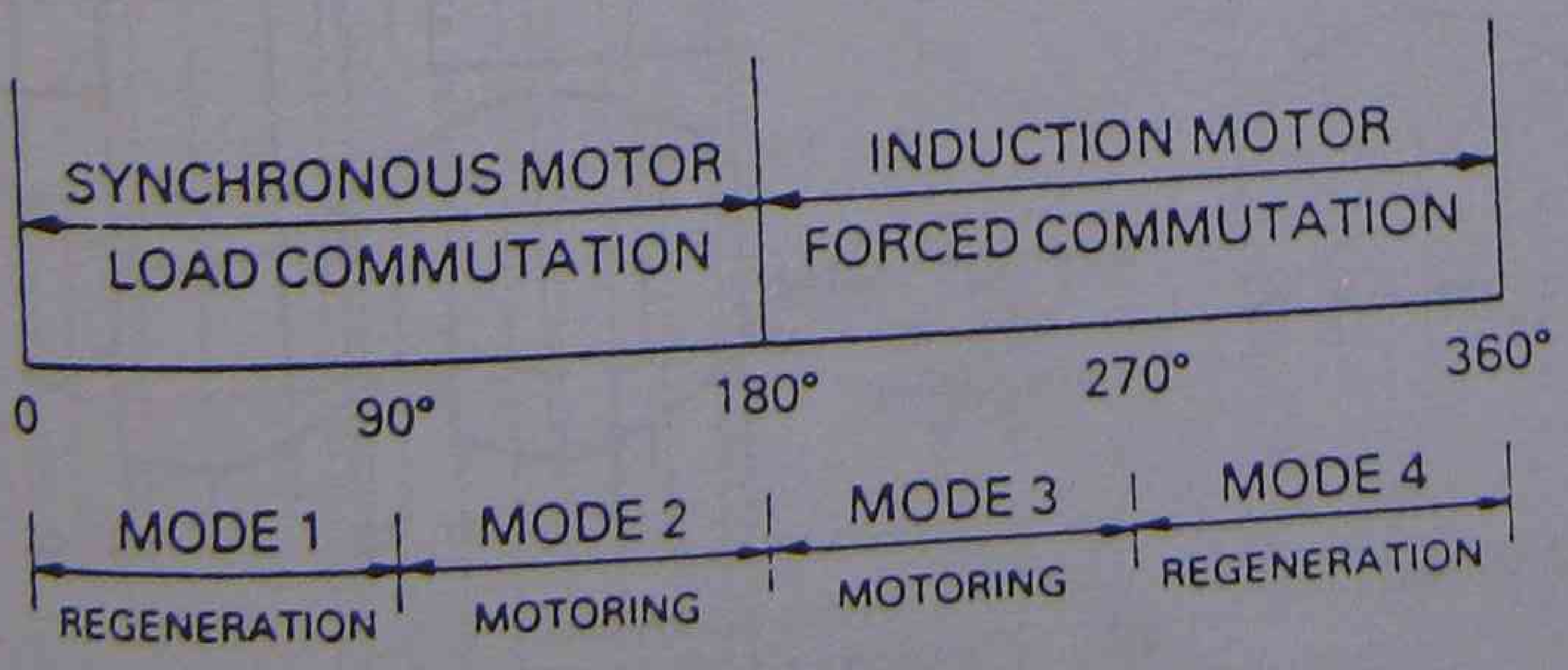
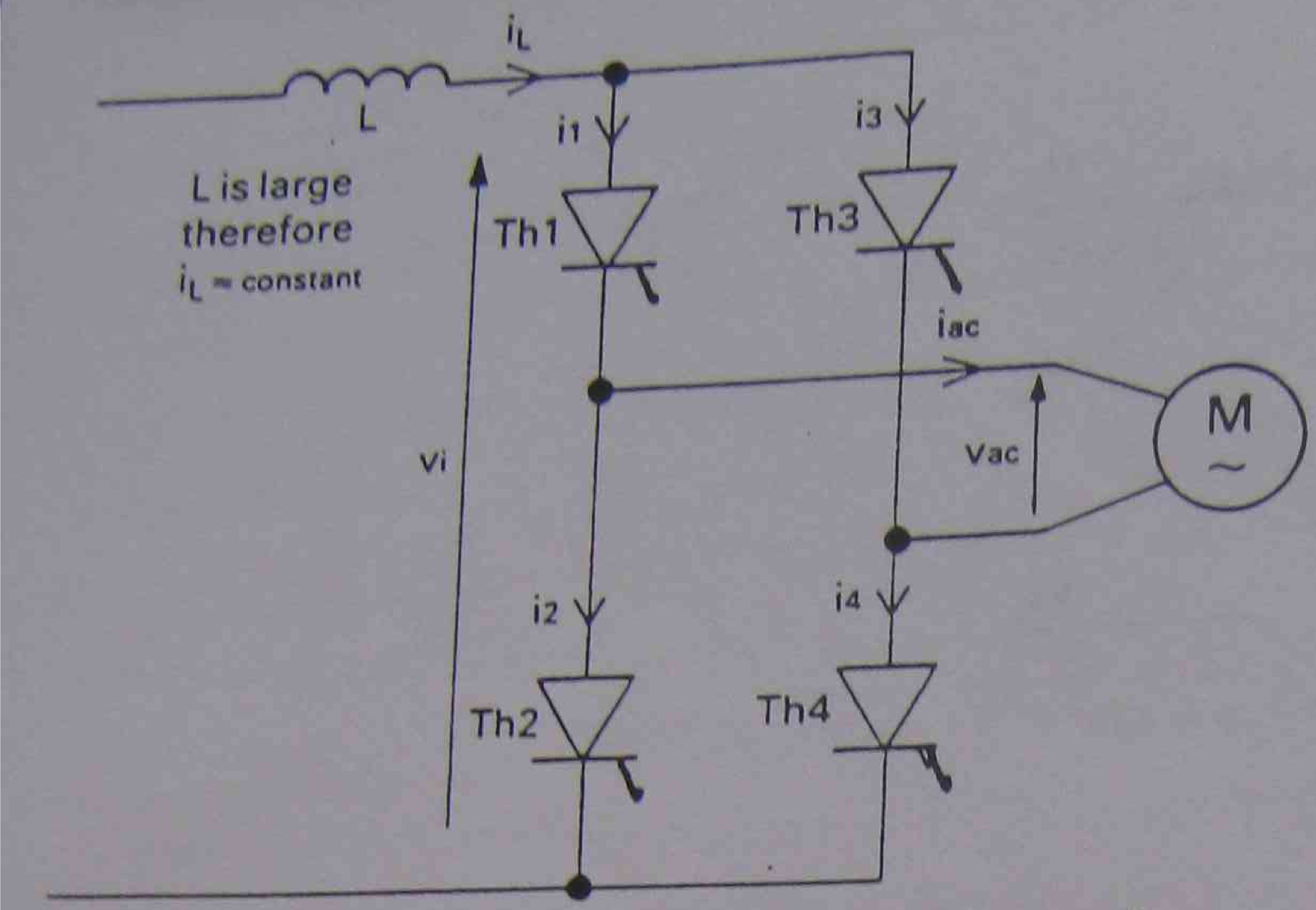


13 Modes of operation with counter-EMF load:

- a $0° \leq \alpha \leq 90°$ = load-commutated rectifier;
- b $90° \leq \alpha \leq 180°$ = load-commutated inverter;
- c $180° \leq \alpha \leq 270°$ = force-commutated inverter;
- d $270° \leq \alpha \leq 360°$ = force-commutated rectifier.

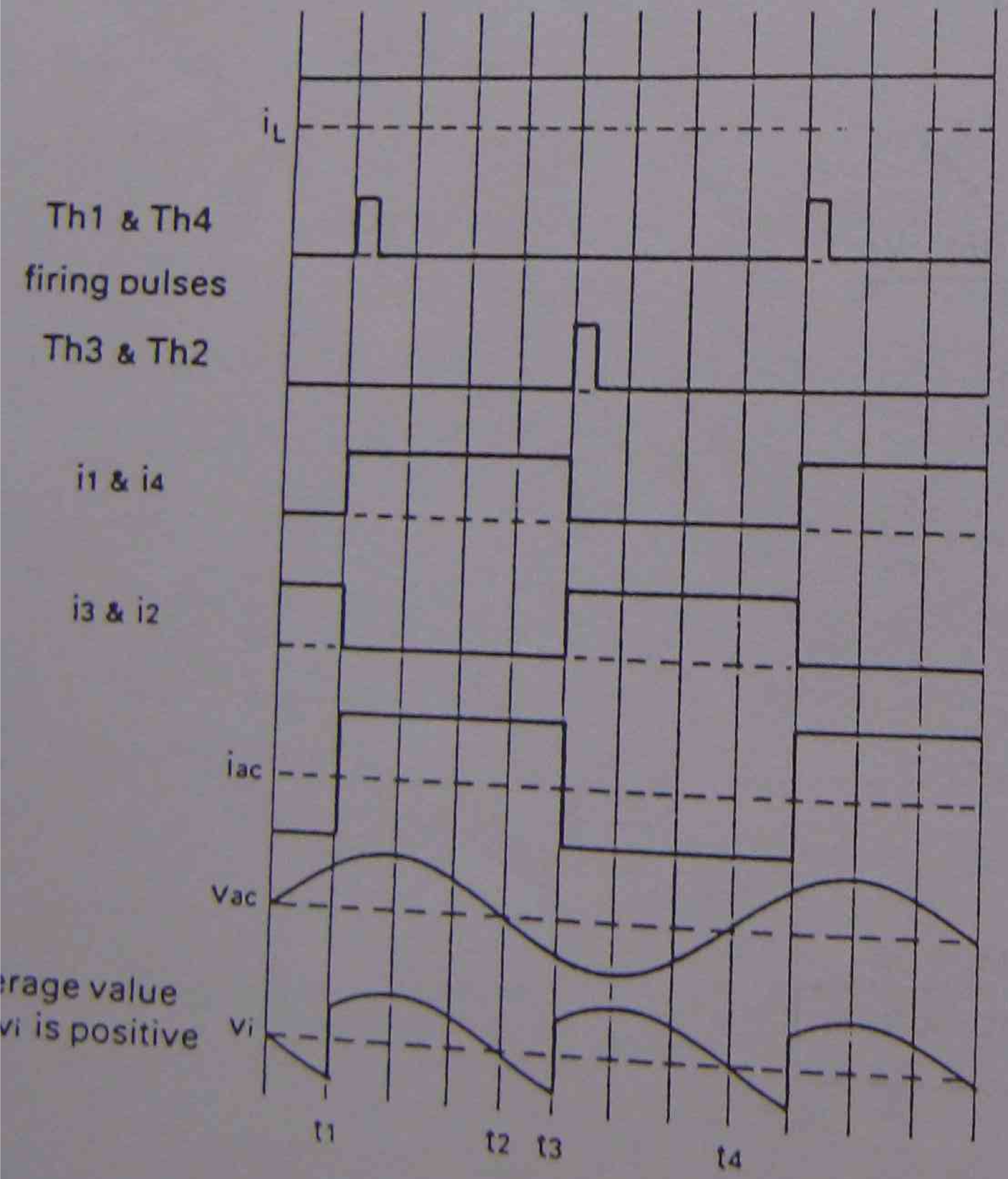


14 Modes of ac machine operation.

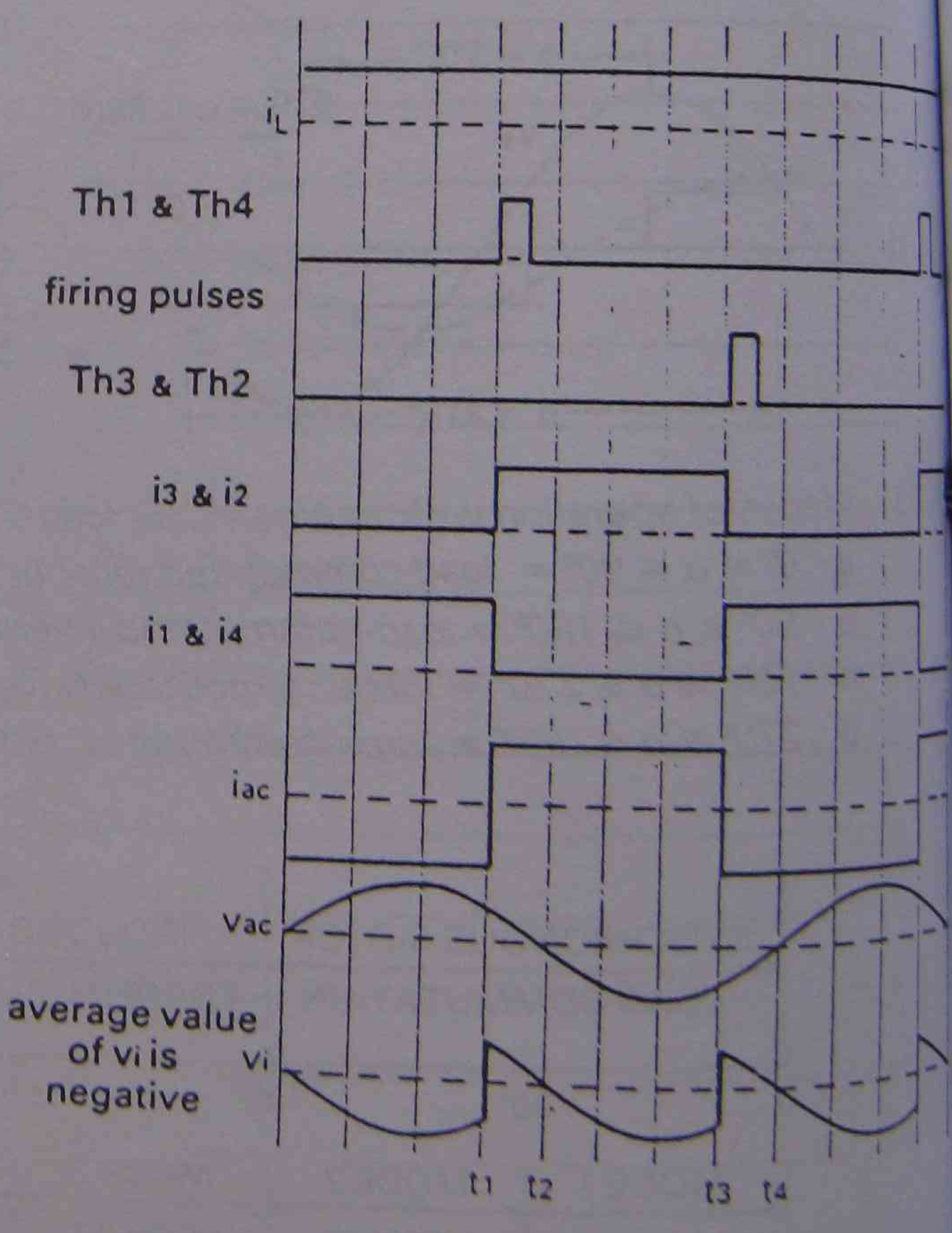


i_L DC link inductor current
iac Motor current
vi Inverter bridge input voltage
vac Fundamental of motor voltage

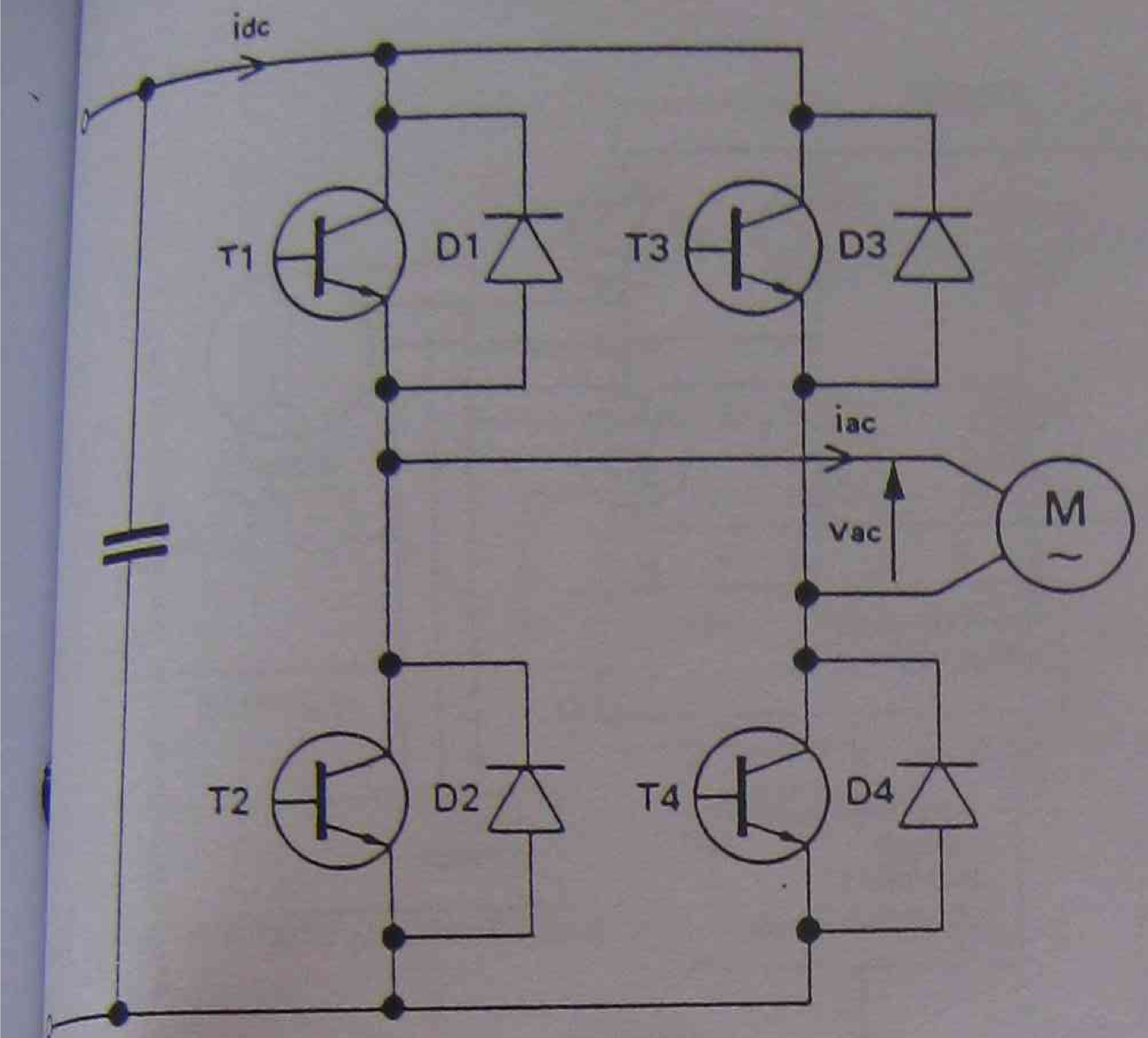
16 *Single-phase current source inverter bridge.*



Current source inverter bridge waveforms when motoring at full load.

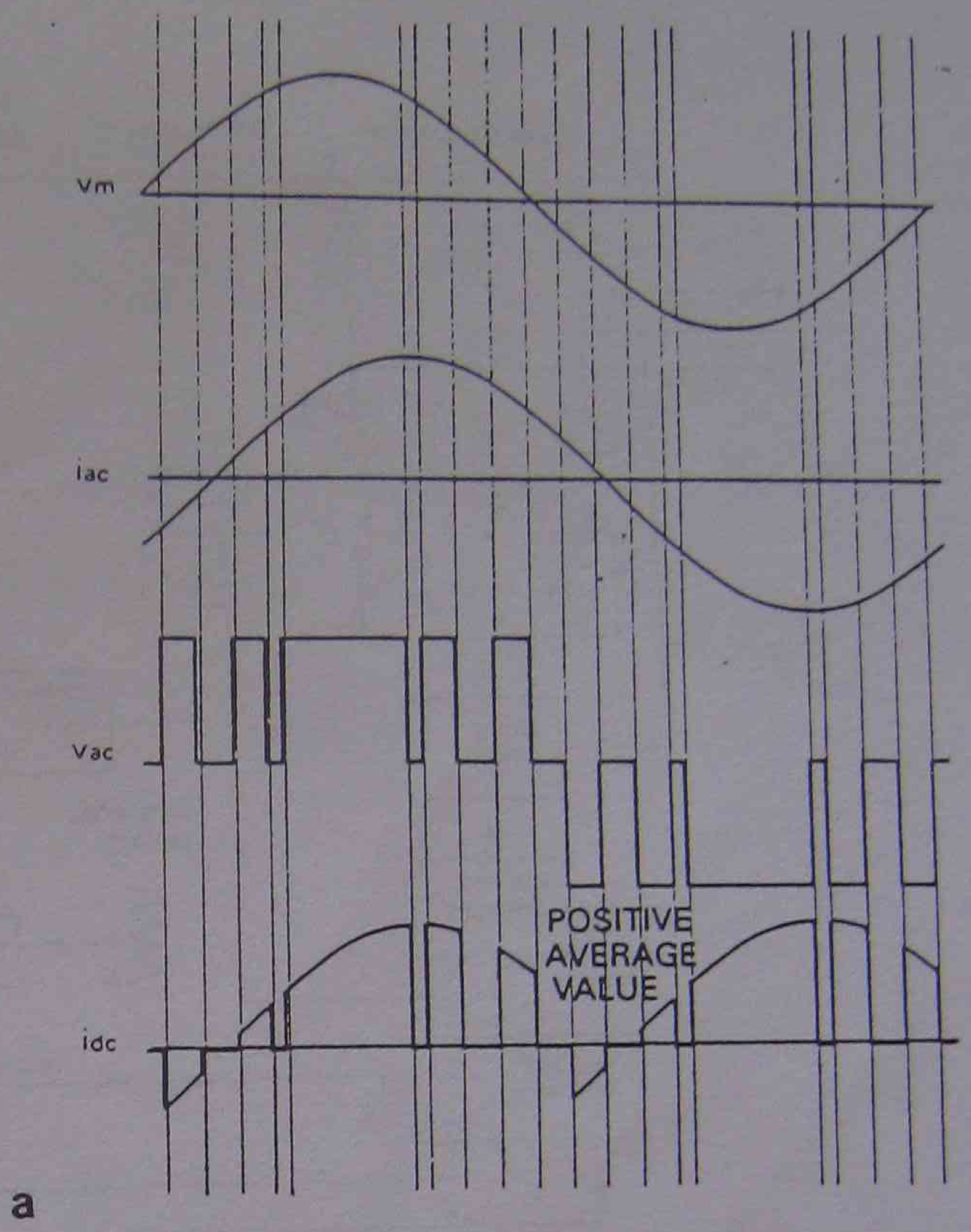


18 *Current source inverter bridge waveforms when the motor is regenerating.*

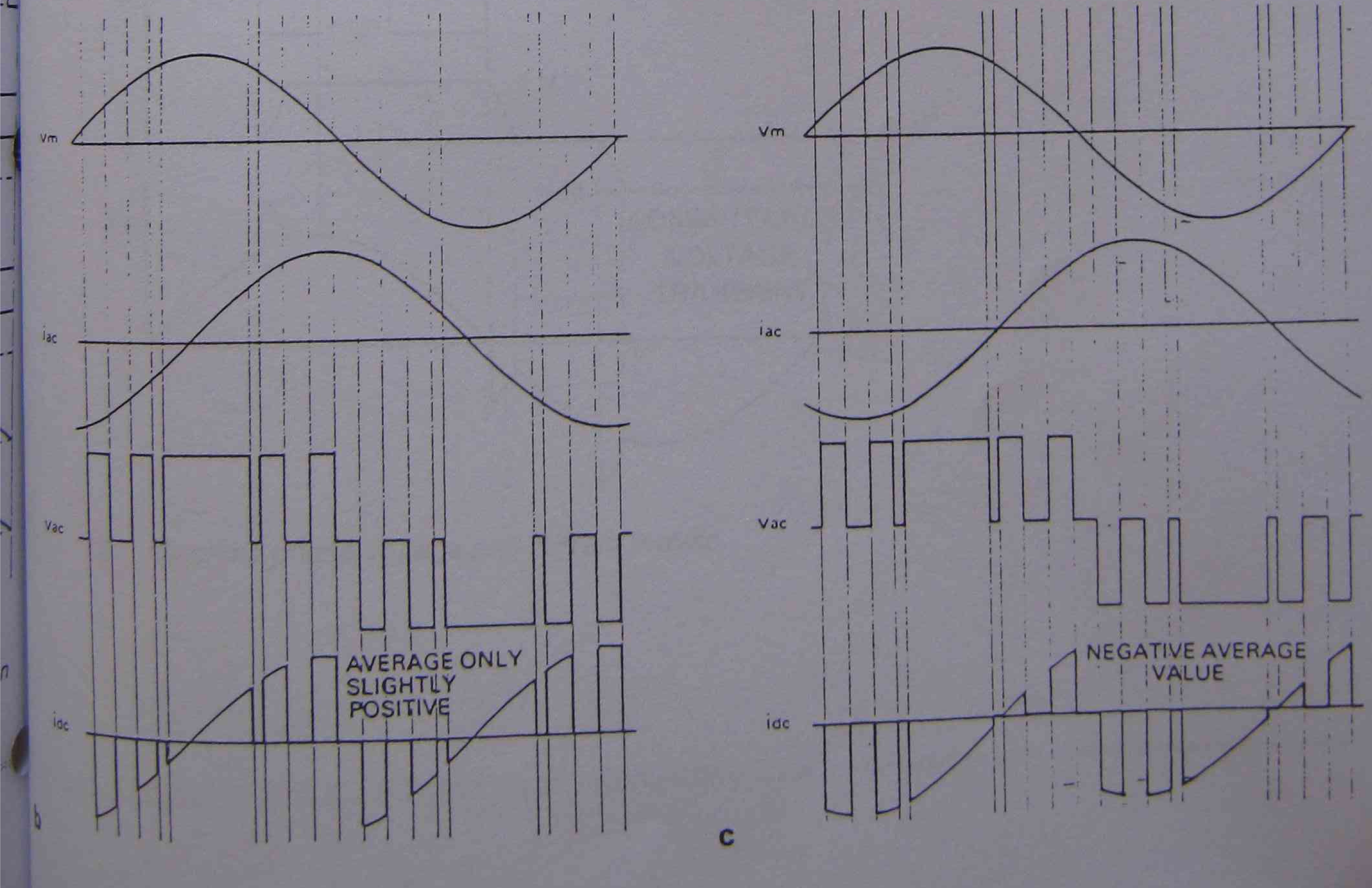


vac Actual PWM voltage applied at motor terminals
Vmf Fundamental of PWM voltage
iac Idealised motor current without ripple
idc DC bus current to inverter bridge

Single-phase PWM inverter bridge.



7 *Inverter bridge waveforms:* **a** *Motoring, full load;* **b** *Motoring, light load;* **c** *Regenerating.*



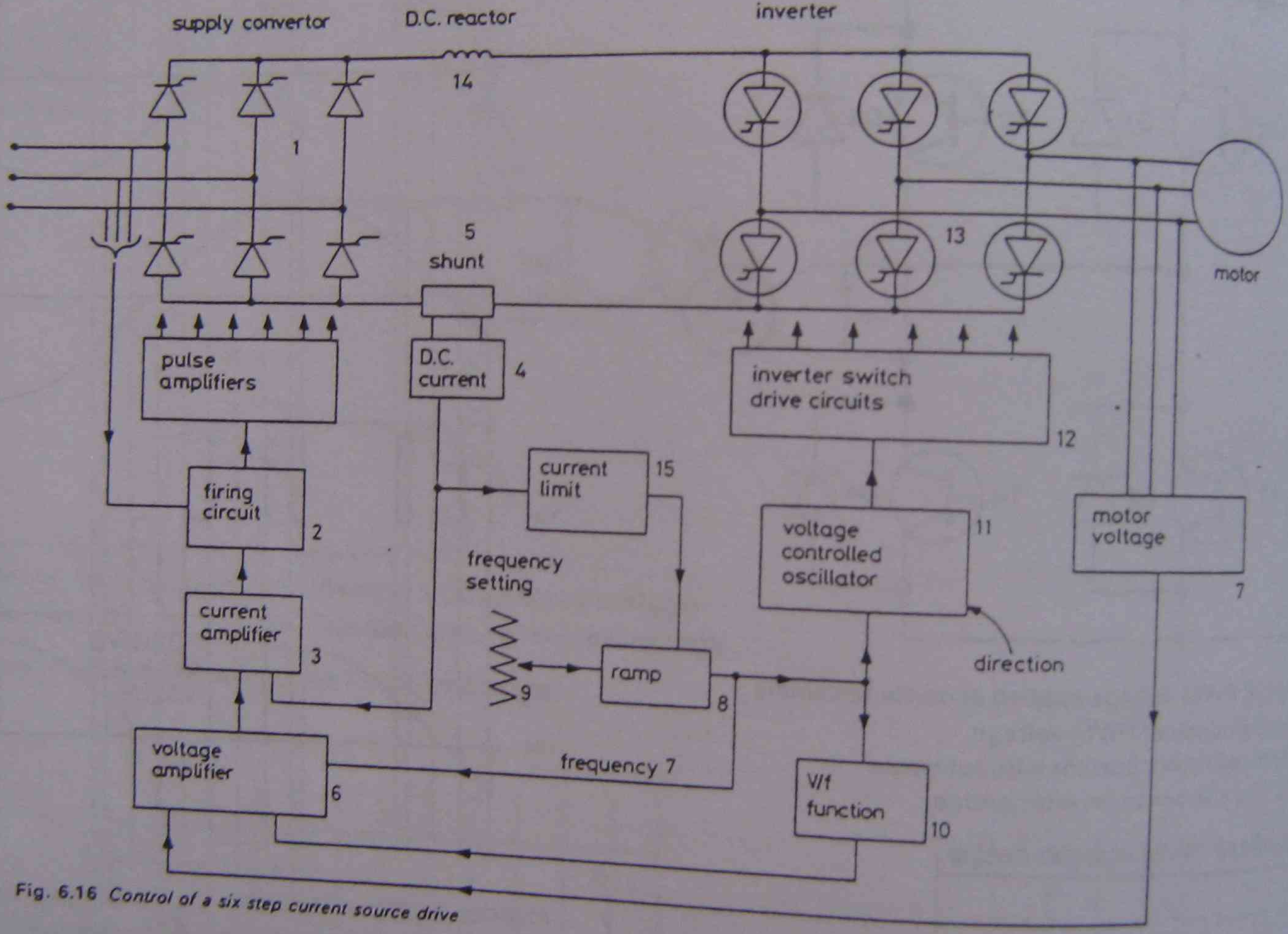
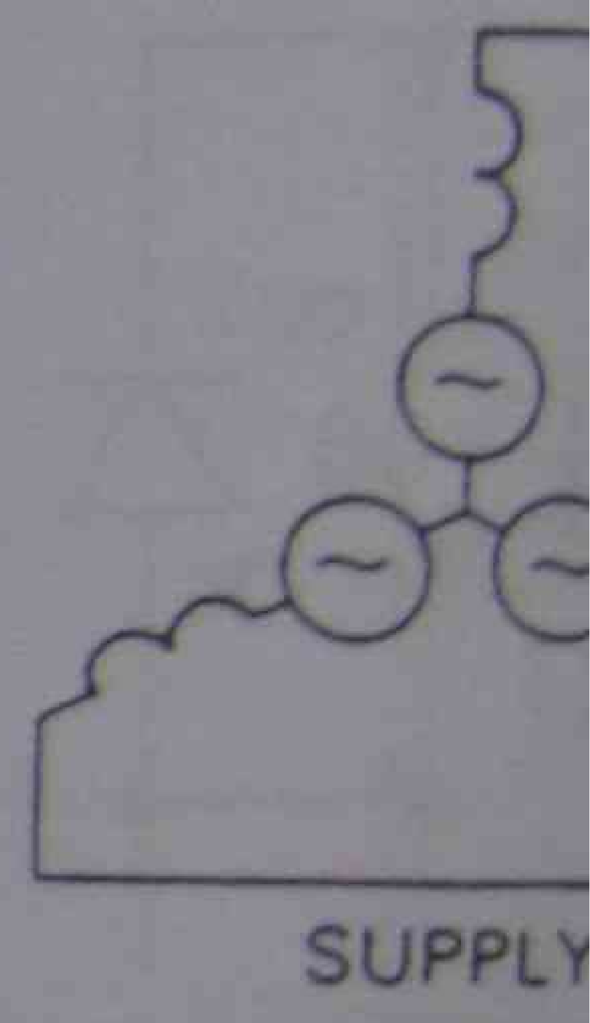
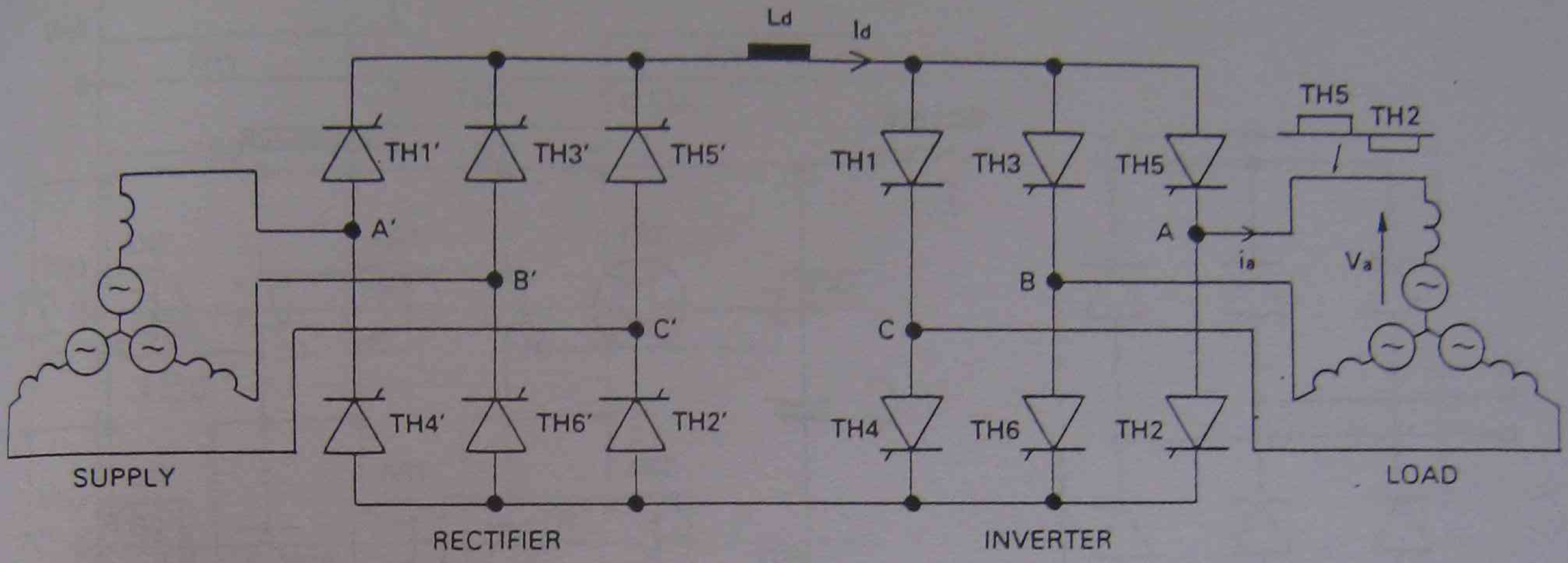


Fig. 6.16 *Control of a six step current source drive*

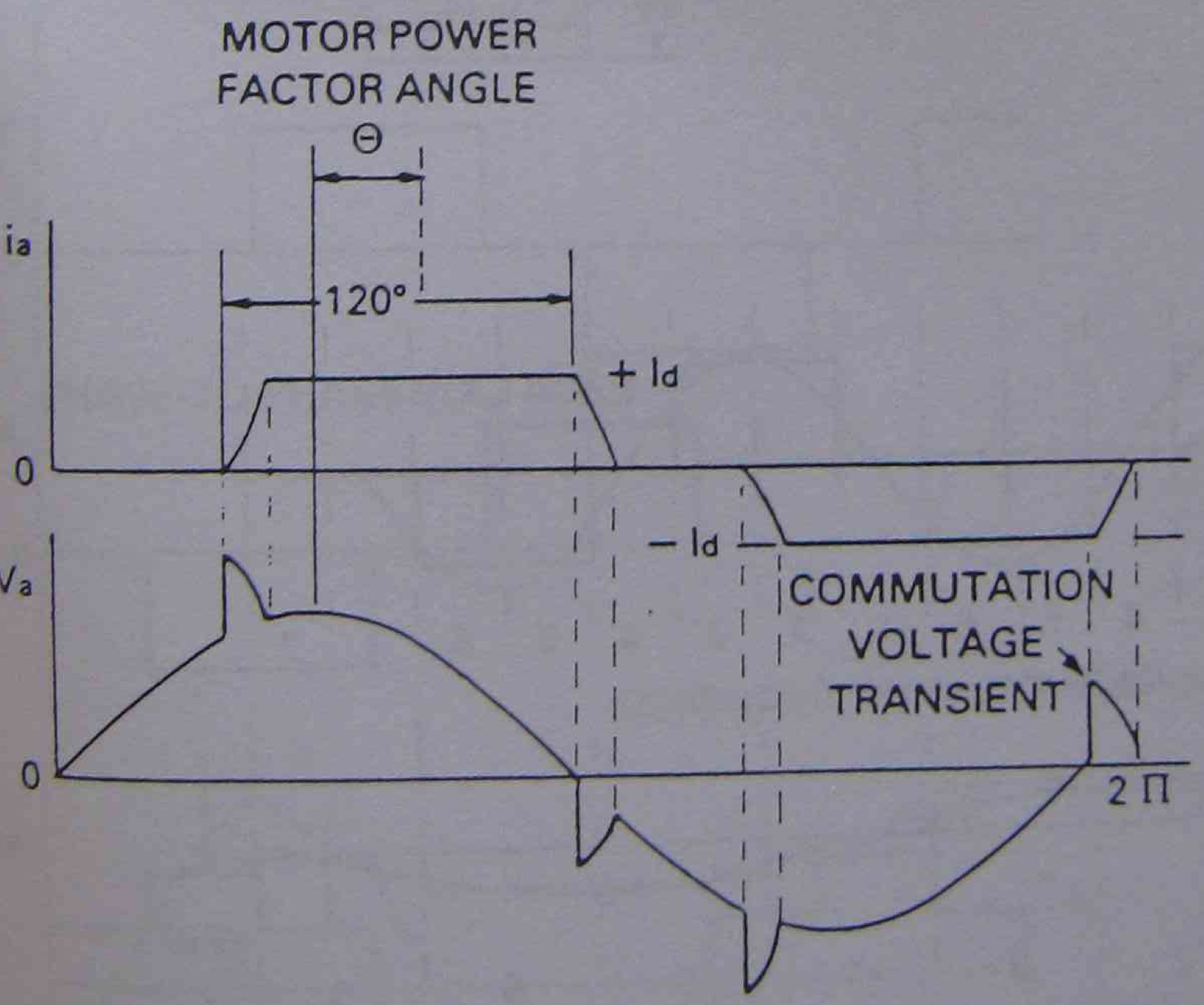
The six step current source inverter drive



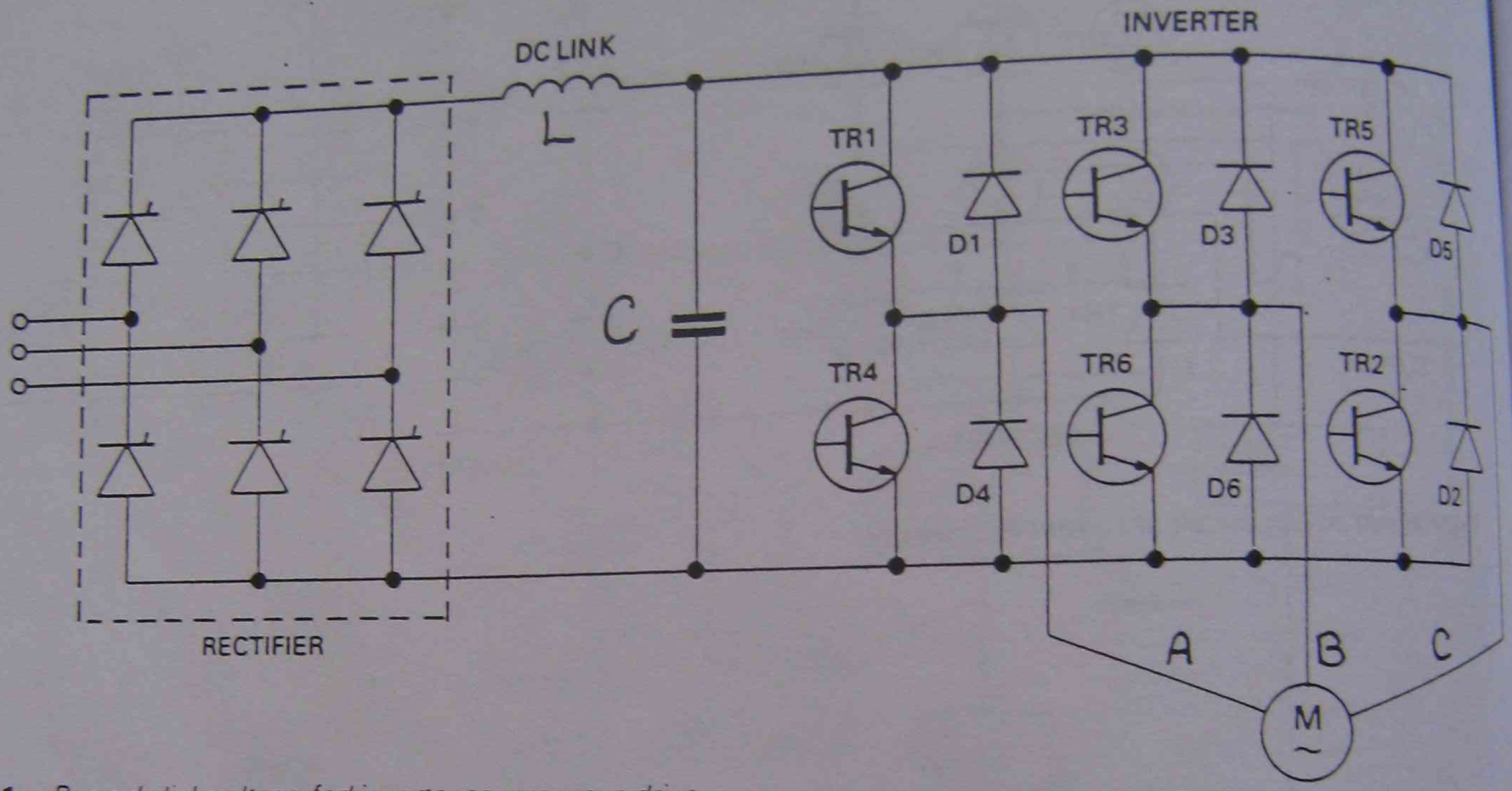
11 *Generalised*



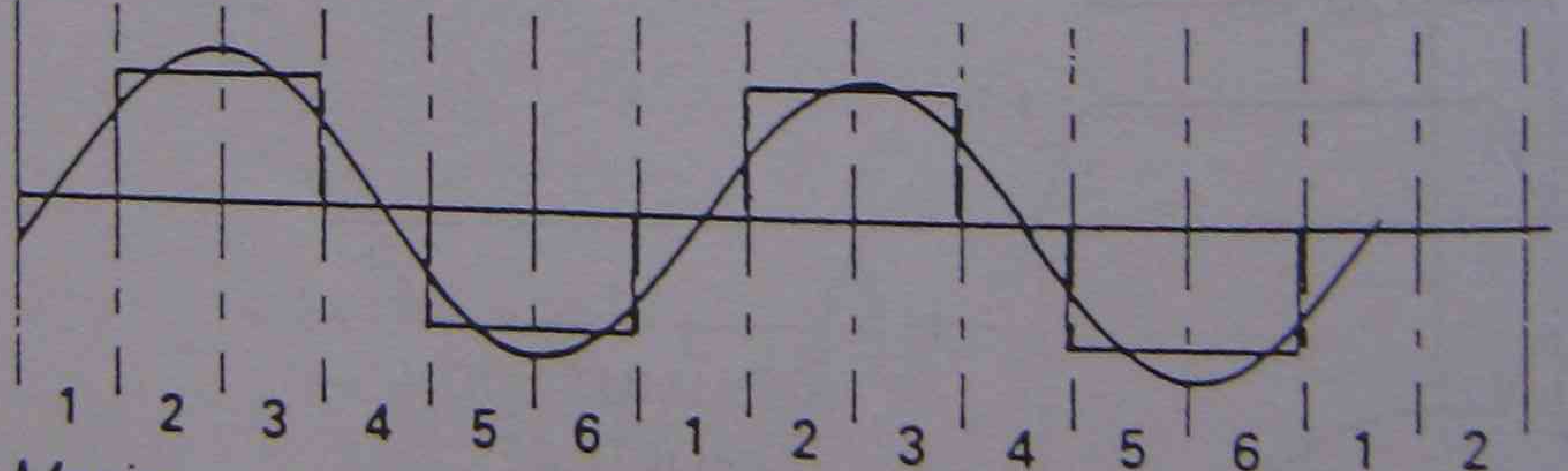
11 *Generalised power circuit of current-fed inverter.*



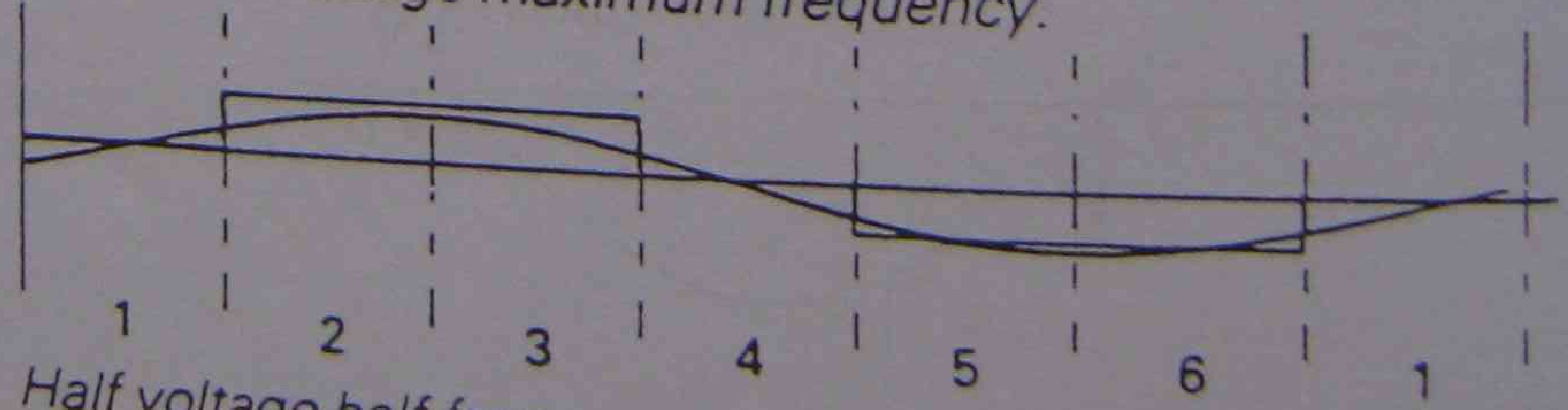
12 *Machine phase voltage and current waves.*



1 *Basic dc link voltage-fed invertor square wave drive.*

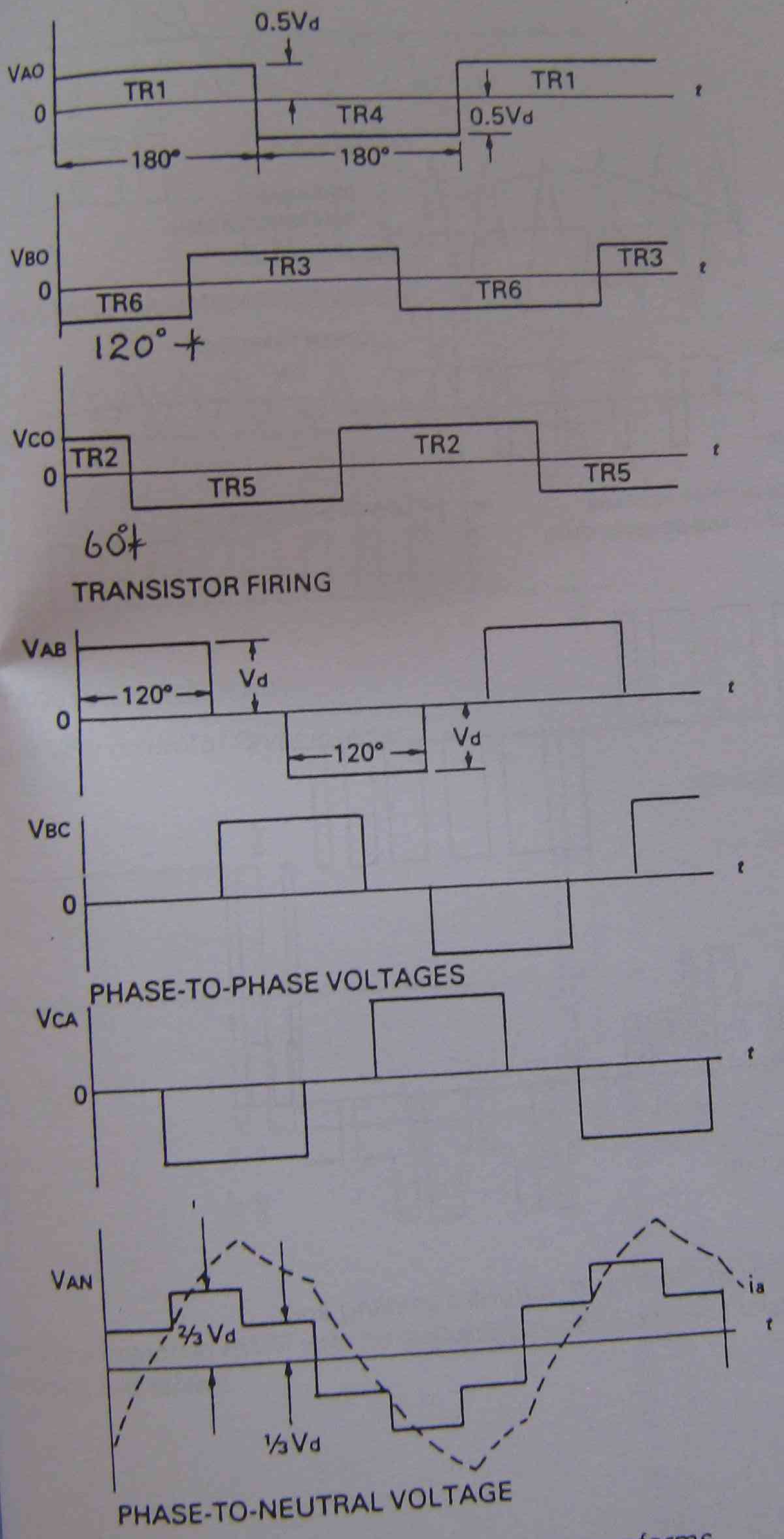


Maximum voltage maximum frequency.

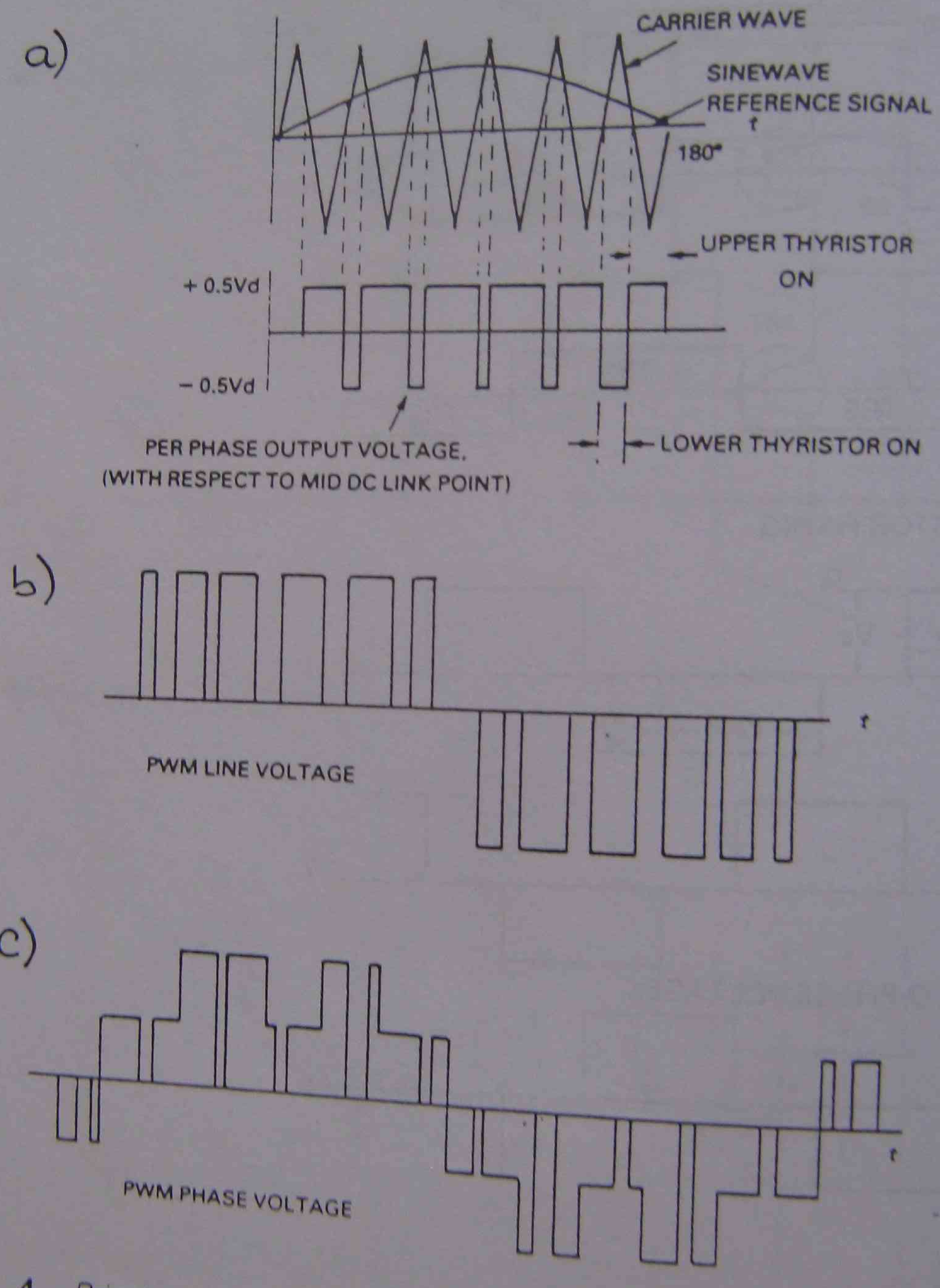


Half voltage half frequency.

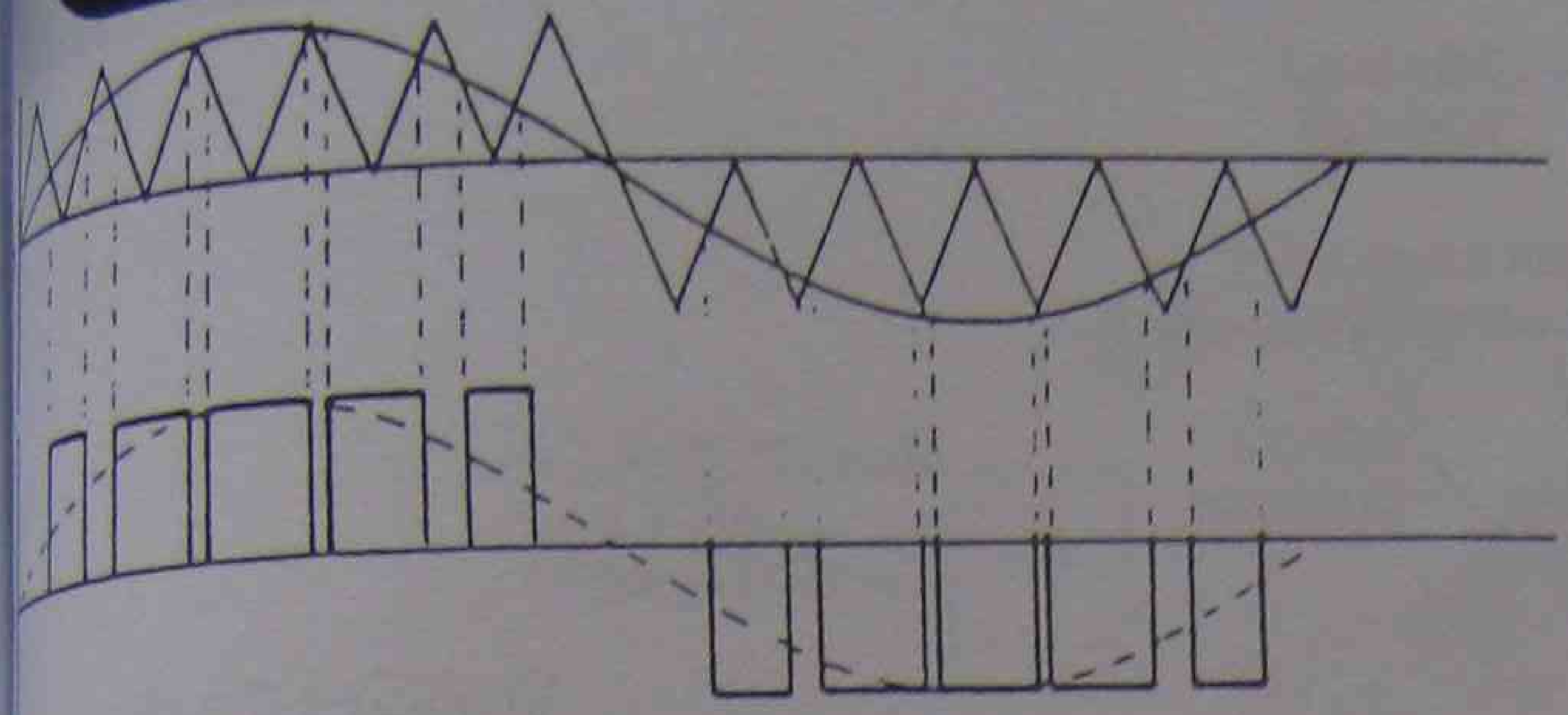
3 *Constant volt/Hz control of square sine wave inverter.*



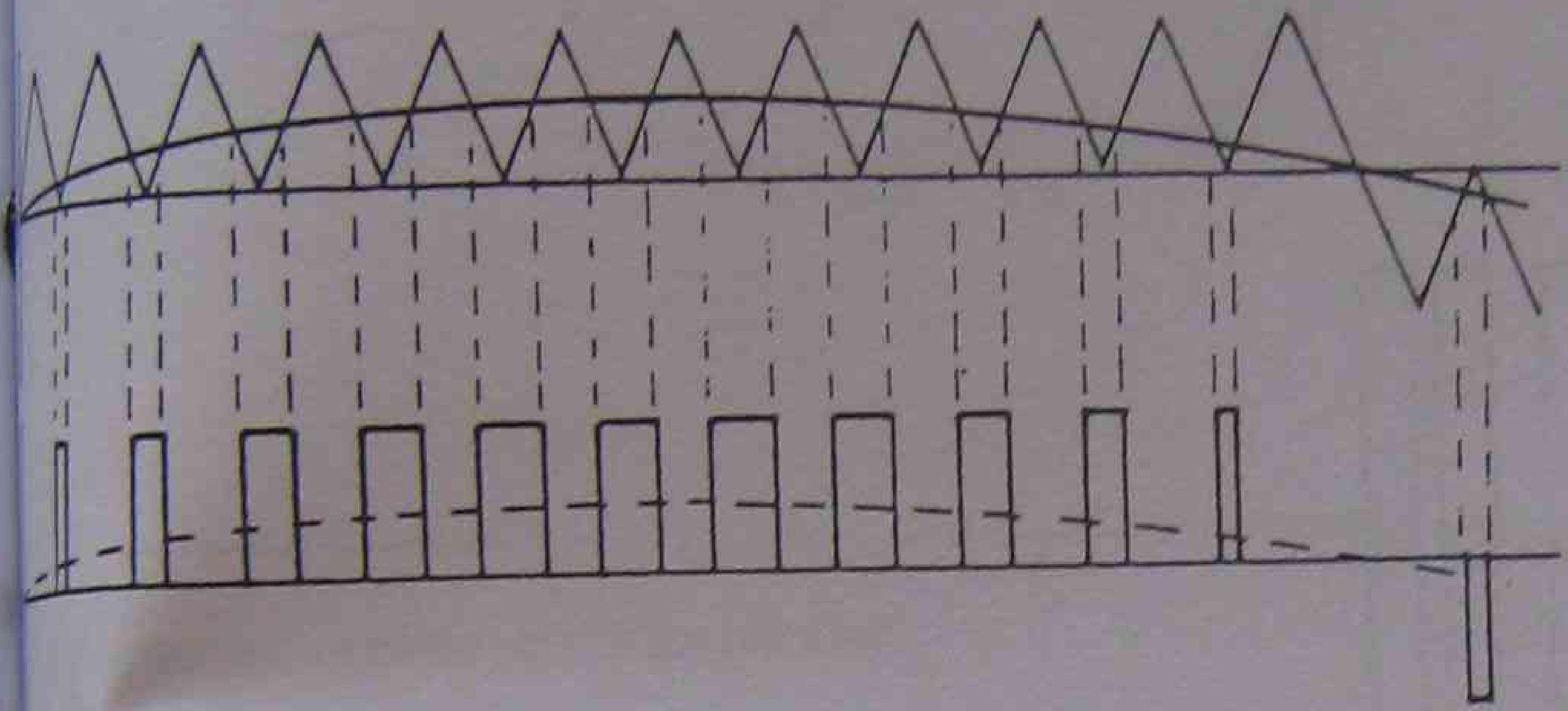
2 *Synthesis of inverter voltage output waveforms*



4 *Principle of sinusoidal PWM with natural sampling and the resulting output line and phase voltages.*

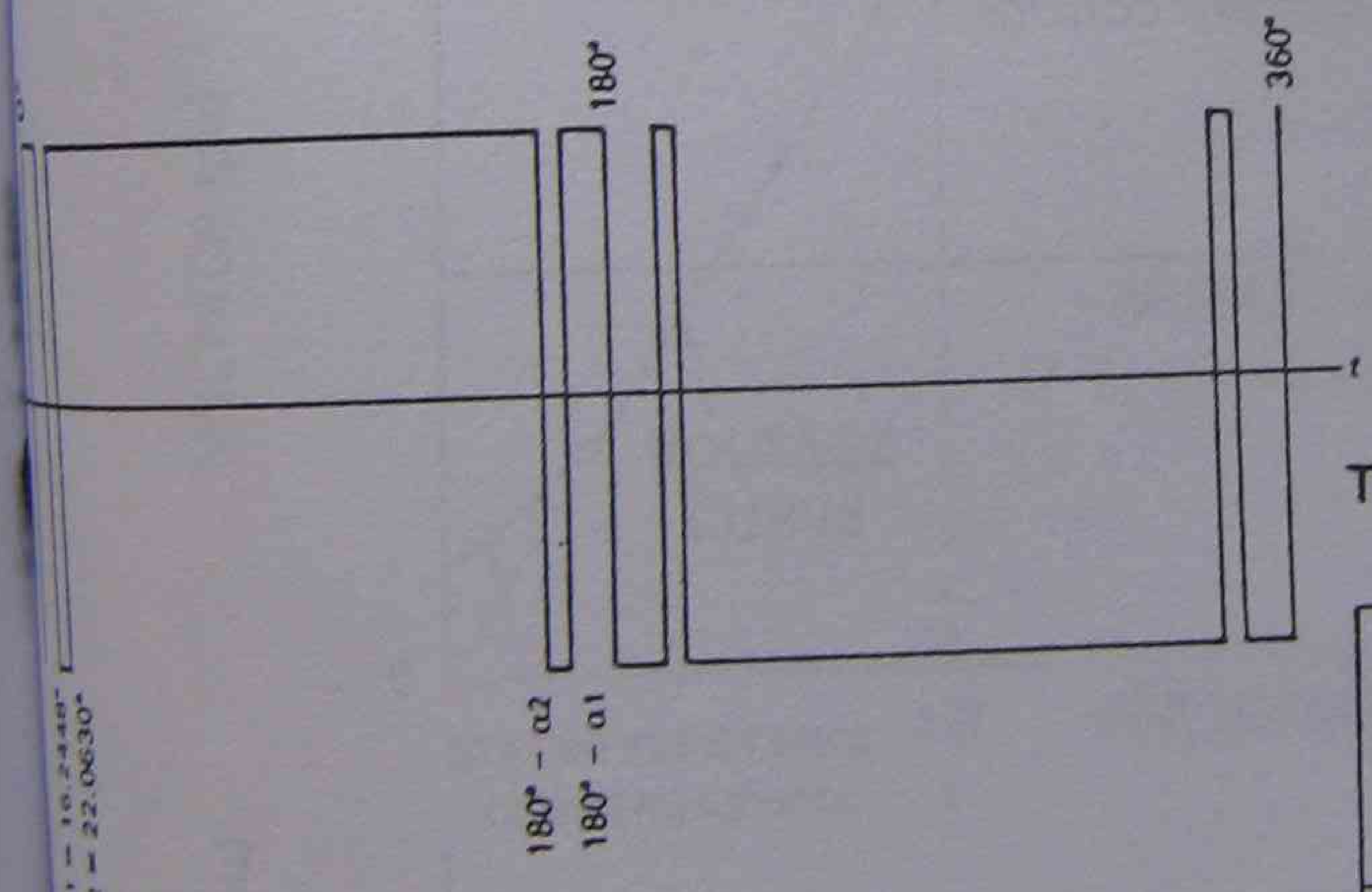


XIMUM VOLTAGE MAXIMUM FREQUENCY



VOLTAGE HALF FREQUENCY

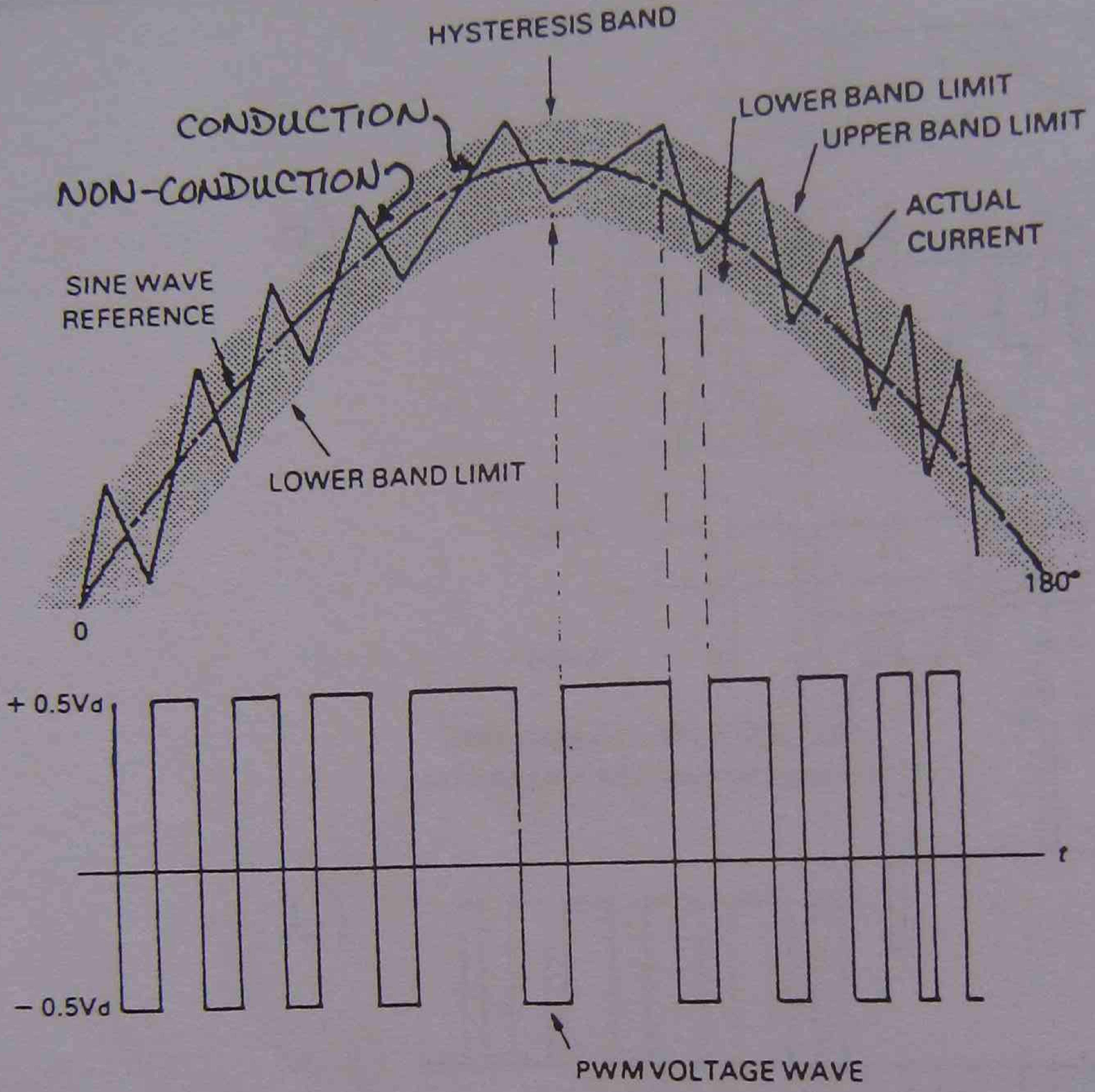
onstant V/Hz control of PWM inverter



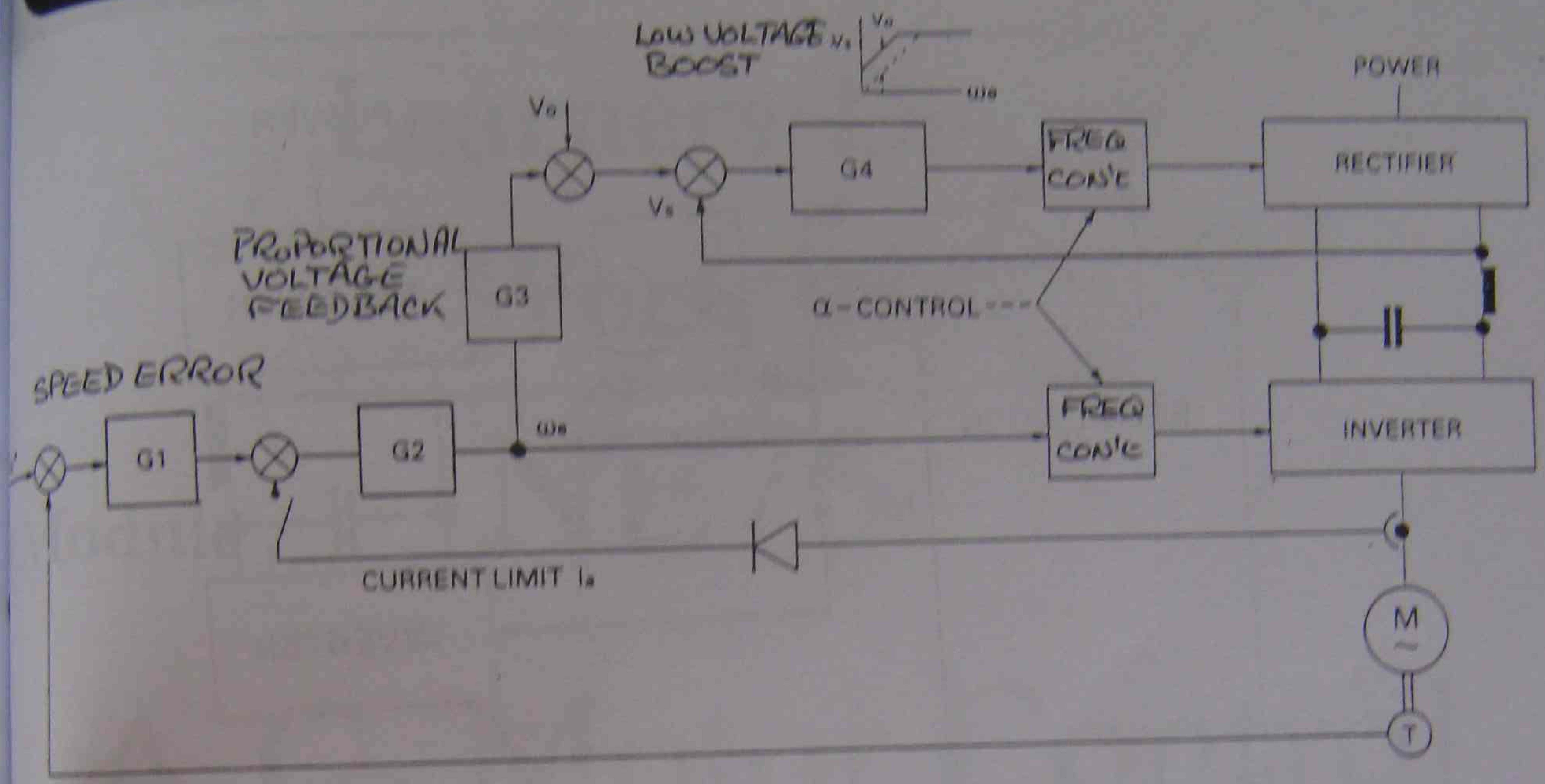
armonic elimination PWM with no fifth and seventh armonics, see table 1.

TABLE 1 Analysis of waveform in Fig.5 showing estimation of 5th and 7th Harmonic.

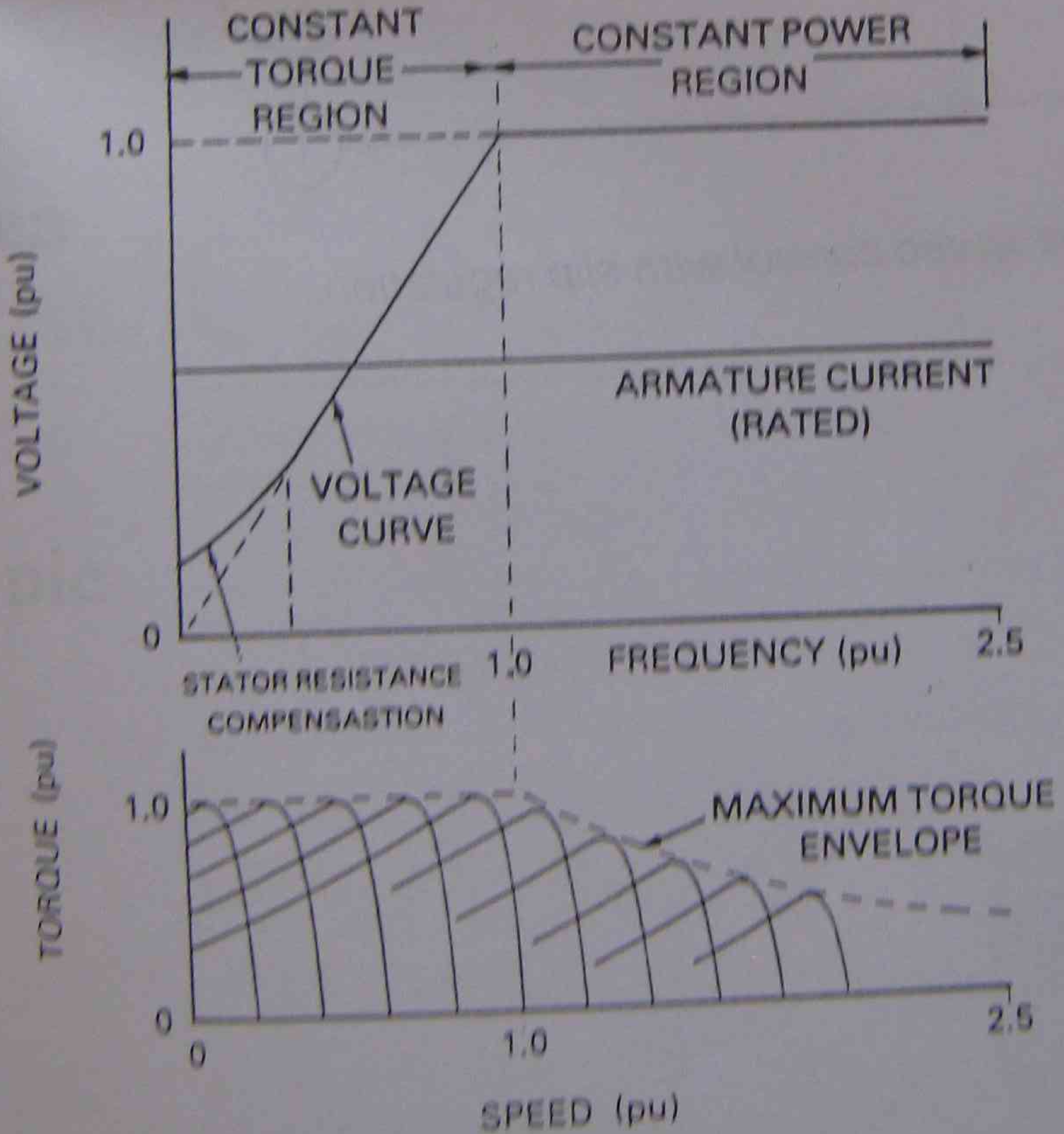
Order of Harmonic	Absolute value of Harmonic - Coefficient	Absolute value of Harmonic as Percentage of Fundamental
1	1.1897	100
3	0.2070	17.43
5	0	0
7	0.0001	0.01
9	0.1086	9.14
11	0.2421	20.31
13	0.3223	27.13
17	0.2030	17.09
19	0.0514	4.33
21	0.0825	6.94



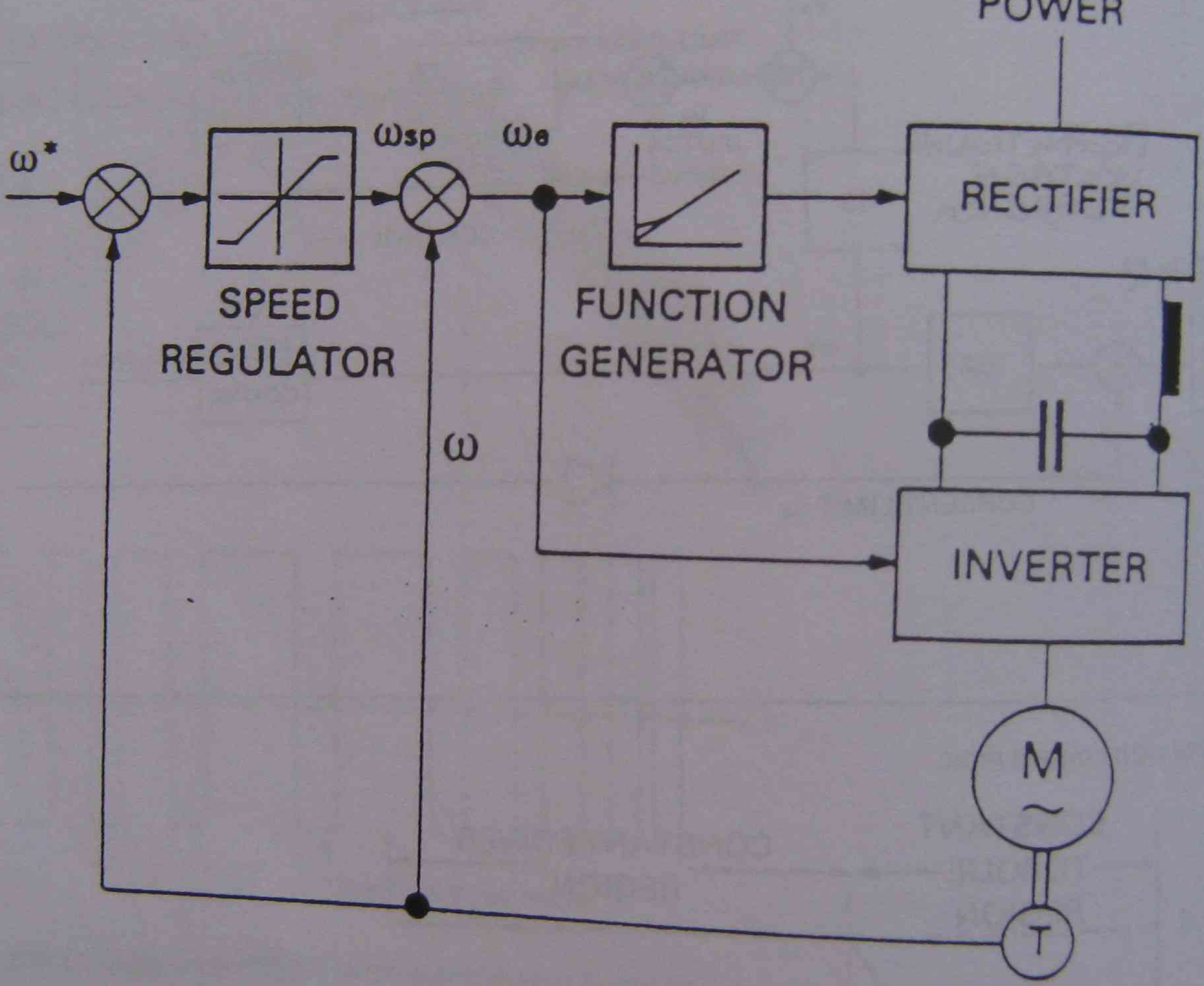
7 *Principle of adaptive PWM with bang-bang current control.*



V/Hz speed control with current limit.



9 *Torque-speed curves of induction motor with variable frequency power supply.*



10 *Constant V/Hz speed control with slip regulation.*

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Module NE76

A.C. Motor Control

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Topic

Session No. 7

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DRIVE SYSTEM	Speed Ratio	Braking Torque	MOTOR DE-RATING Continuous Use at:		Torque Response	SPEED HOLDING 0-100% Load Change	Speed Linearity	Starting Torque	CONNECTIONS Controller to Motor	SUMMARY	
			Full Load	Low Speed						Principal Advantages	Principal Disadvantages
DC Motor + Converter	1:20 (Analogue AVF) 1:100 (Analogue dc tacho) 1:400 (Temp compen-sated dc tacho or encoder) 1:1000 (Encoder with digital lock)	150% (with 4Q drive)	Dependent on motor rating and 6, 3 or 2-pulse bridge: no de-rating above 10kW with 6-pulse	As full load if forced air cooling. 1:5 speed ratio at 90% nominal torque without forced cooling	<10ms with 6-pulse bridge, no choke in motor supply	2·0%; AVF with IR compensation. 0·1%: Anal.tacho. 0·05%: Digital encoder	Typically 0·1% With digital speed feedback <0·01%	150% 20-30s	AVF: 4 Closed-loop: 6	● Wide speed range ● Good starting torque ● High efficiency ● Economic cost of controller ● Digital communication possible ● Braking facility	● Expensive motor ● Brushgear maintenance ● Possible commutation failure on loss/dip of supply ● Torque variation at low loading ● Low/part loads reduce brush life ● Low protection class (motor)
AC Motor + Inverter (VVVF or PWM)	1:20	approx 100% for voltage-driven inverters. 100% (regen) for PWM. DC braking 50-100% depending on motor: – non-linear; also high losses	5-15% dependent on quality of sine waveform output	Dependent on speed range and motor/inverter characteristics. 90% nominal torque over full speed range with forced cooling. 1:4 speed ratio at 90% nominal torque without forced cooling.	Dependent only on torque response of motor	1-2% with slip compensation	Typically 0·1%	Depends on quality of sine waveform output: for 150% FLC 20-30s, torque 75-150%	Open loop: 3	● Maintenance economy ● High protection class ● Motor availability ex-stock ● Simple application and set-up ● Economic cost of motor ● Digital communication possible ● Overload and fault protection ● Speed range to >100% nominal ● Braking facility	● Possible motor instability at low speed/ light load ● Possible acoustic noise ● Average speed range available
AC Motor + Flux Vector	1:400	150% dynamic or regen	5%	Forced cooling is usual: typically 95% nominal torque over full range	<1·0ms	0·05%	<0·01%	150% 30s	Power: 3 Encoder: 4-6	● Excellent torque response ● High starting torque ● Widest range of applications ● No zero-torque deadband ● Excellent stability ● Maintenance economy ● Overload and fault protection ● Speed range to >100% nominal ● High protection class ● Braking facility	● Additional converter required for regeneration into supply
AC Switched Reluctance Drive	1:100	Effectively max 20% braking torque	Nil – Special motor	Nil – Special motor	Typically 40 ms	0·01%	0·3%	150% 5s	Power: 6 Feedback: 4	● High efficiency ● High standard speed range ● Overload and fault protection ● High protection class	● Specialised motor design ● Applications limited ● Cogging tendency at low speed ● Acoustic noise ● 6 power cables required for motor
AC Eddy-Current Coupling	1:20 1:40 Depending on number of poles	Nil	Nil	Dependent on cooling. Typically lower speeds not available at >50% motor torque.	Not good	0·5%	1·0%	Dependent on coupling frame size. 150% possible	Power: 2 Feedback: 2	● Economic maintenance ● Good starting torque	● Efficiency low at low speed ● Poor torque response ● No braking facility

A.C. Motor Control.

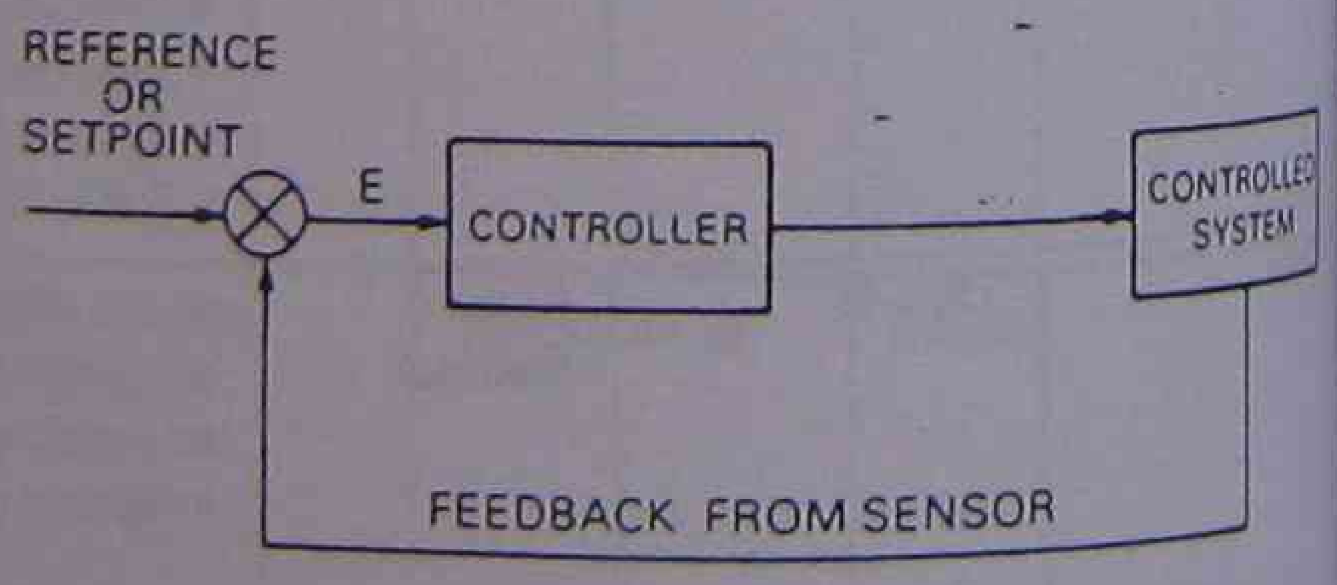
Topic -

4 CLOSED-LOOP CONTROL

The general principle of closed-loop control is illustrated in Fig.1. The principle applies not only to variable-speed drives but also to other electrical, mechanical, hydraulic or pneumatic systems.

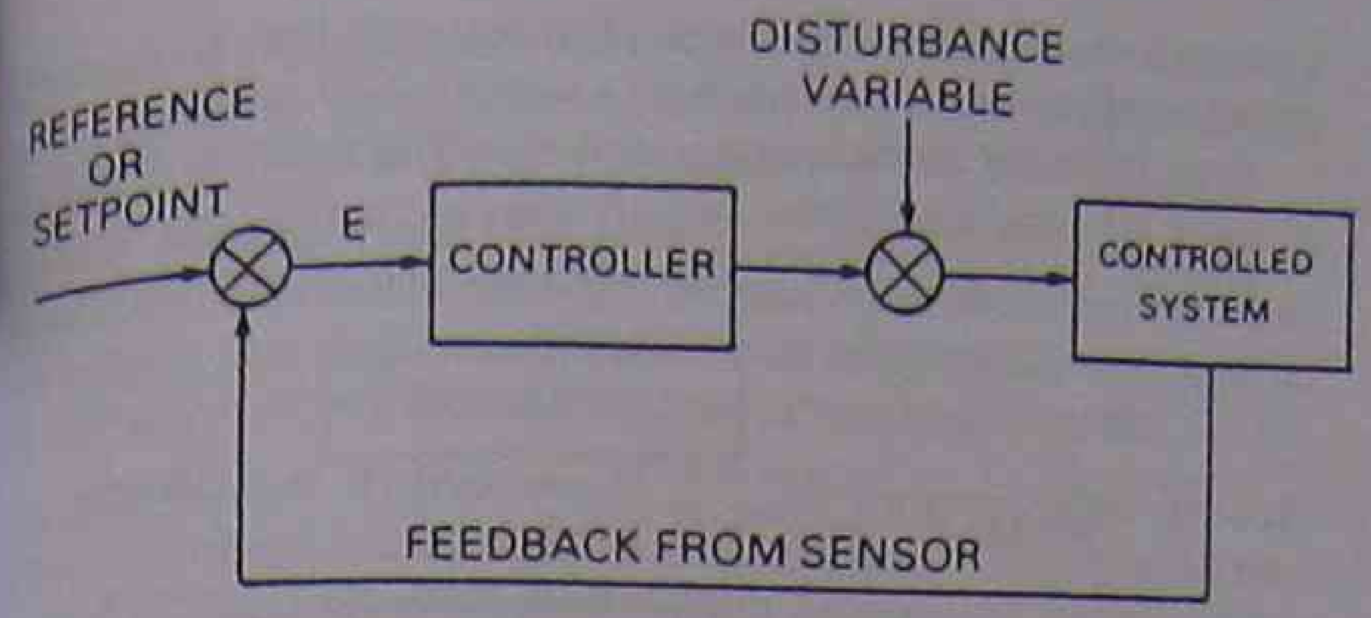
The closed-loop system is characterised by a feedback signal, derived from a sensor in the controlled system. This signal monitors the actual behaviour of the system, and is compared with (or subtracted from) the reference signal. The magnitude and polarity of the resulting error signal, E, are directly related to the difference between required and actual values of the controlled variable — which may be the speed of a motor, temperature of a furnace, level of liquid in a tank, and so on. The error signal is amplified by the controller, and the controller output makes a correction to the controlled system, reducing the error signal.

As the error signal reduces, it eventually reaches a value at which the control system gain is insufficient to make any further correction to the controlled system, resulting in a band either side of the set point, known as 'dead space' in which no effective regulation occurs. Increasing the control system gain reduces the width of the dead space, but increases the tendency of the overall system to become unstable, ie to break into oscillation. The time-constant or rate of response of the system introduces a delay between input and output. If the delay is such that the output of the system is in antiphase with the input, the feedback will add to the error signal rather than reducing it. If the system gain is sufficiently high, oscillation instability will result. Obviously this is undesirable in a control system, and the gain must therefore be limited in practice.



E = ERROR SIGNAL

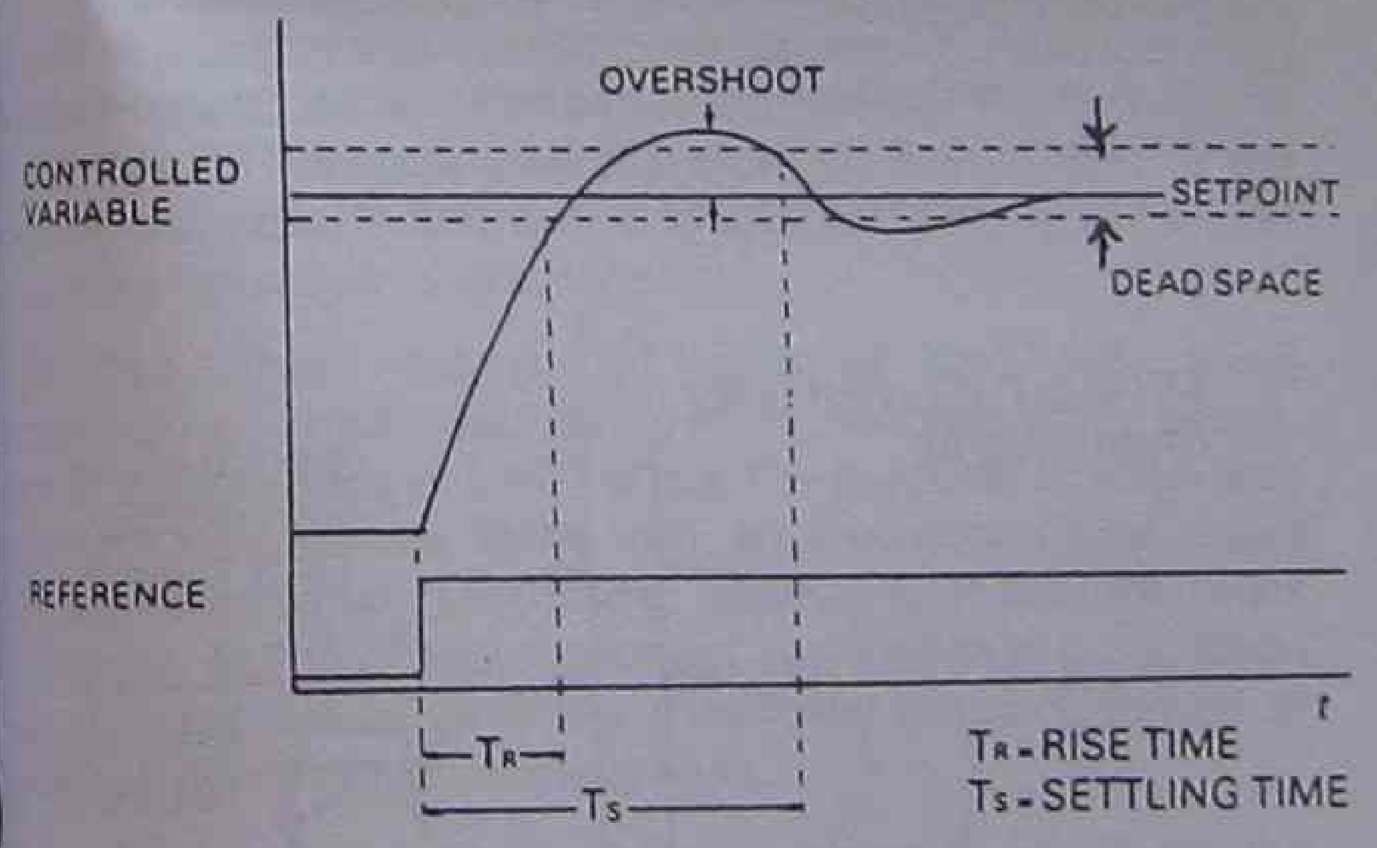
1 *Closed loop control system*



E = ERROR SIGNAL

2 *Closed loop control system with disturbance variable*

A development of the basic closed-loop system is shown in Fig.2, where an external disturbance variable is introduced between controller and controlled system. In speed control systems, the disturbance variable might represent load fluctuation; in temperature control, the heat loss resulting from opening the furnace door; in liquid level control, variations in the rate at which liquid is drawn from the tank. Therefore the controlled system is affected directly by the disturbance variable, causing a change in the feedback signal which results in compensation by the control system.

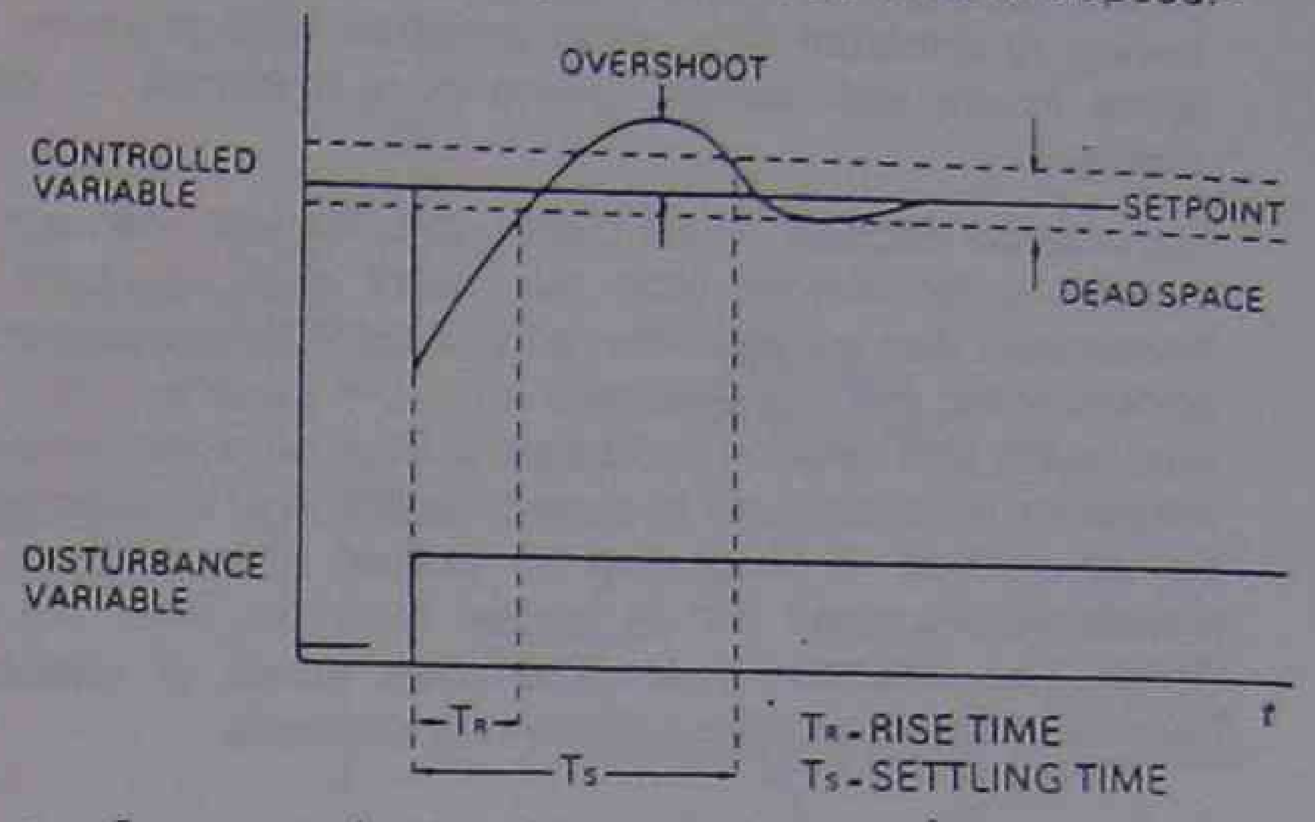


3 *Response of a closed-loop system to a step change in reference*

The response of a closed-loop system to step-changes in reference and in disturbance variable are illustrated in Figs.3 and 4.

The system response shows a characteristic Rise Time, T_R, which is the time that elapses between the step-change in reference or disturbance variable and the initial entry of the controlled variable into the dead space.

Typically, an overshoot then occurs, and when the controlled variable re-enters the dead space to remain within it, the system Settling Time, T_S, is considered to have elapsed.



4 *Response of a closed-loop system to a step change in disturbance variable*

If no overshoot occurs, the system is said to be overdamped, Fig.5a, and will tend to respond in a sluggish manner, whilst multiple overshoots are characteristic of an underdamped system, Fig.5b, which is liable to instability. The ideal is the critically-damped response illustrated in Fig.5c.

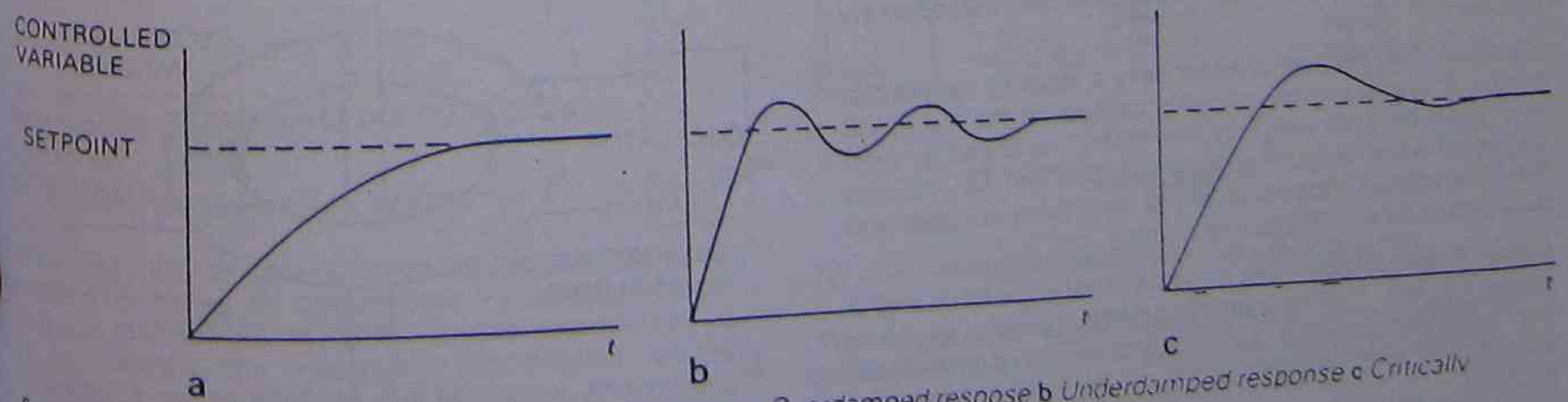
Feedback Sources

In the closed-loop control of motors by means of thyristor drives, the feedback to the control loop is from a transducer or sensor which converts a non-electrical quantity (eg speed) into an electrical quantity (eg a voltage), or converts an electrical quantity to a form which is compatible with the regulator input circuits (eg current to voltage). The following are examples of feedback sources which are commonly encountered.

Tachogenerator. This is a small permanent-magnet dc generator, usually driven directly by the motor shaft via a coupling attached to a stub shaft at the non-drive end. The output is a dc voltage which is proportional to the speed of the motor, and whose polarity is determined by the direction or rotation.

AC Tachogenerator. Mounted directly to the motor in the same way as a dc tachogenerator, the output is an alternating voltage whose magnitude and frequency are proportional to speed. Usually the output is rectified to give a dc signal. The ac tachogenerator is unsuitable for reversing applications, since its output is unipolar. Operation at low speeds is often unsatisfactory owing to the high ripple content of the signal.

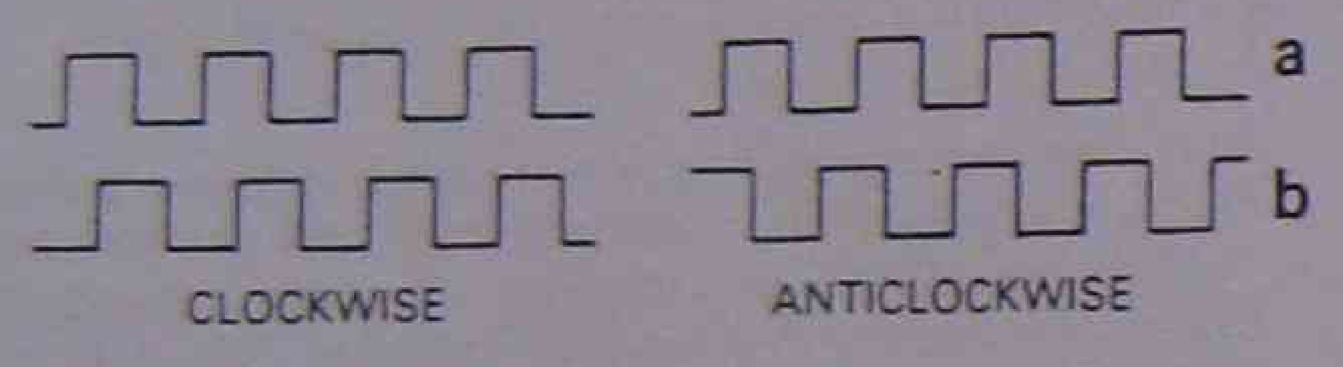
DC Voltage Transformer (DCVT). This device is used to provide an electrically-isolated armature voltage feedback



5 *Effect of system damping on response to a step input:* **a** *Overdamped respose* **b** *Underdamped response* **c** *Critically damped response*

signal as an alternative to non-isolated potential dividers. An internal oscillator modulates the primary voltage (motor armature voltage) impressing upon it an ac component which enables normal transformer action to take place. The secondary voltage, rectified and smoothed, becomes the feedback signal. Bipolar DCVTs are available for reversing applications, but armature voltage feedback is only suitable where accuracy of speed control is not critical, because (owing to armature resistance) armature voltage varies to some extent with current, giving poor response to load changes.

Incremental Encoder (Pulse Generator). The encoder, again usually driven directly from the motor shaft, contains a transparent disc marked with radial lines. A light source and photoelectric cell are arranged near the periphery of the disc, such that rotation produces a train of pulses whose frequency is proportional to speed. Direction of rotation can be determined if necessary by pulses from a second photocell, displaced 90° in phase from the first, Fig.6. Quadrature detection logic determines sense of rotation from the phase relationship of the two channels.



6 *Pulse trains corresponding to bidirectional rotation of an encoder.*

The encoder has the advantage that its output is a pulse train whose frequency is not affected by temperature or attenuation of long cable runs as is the analogue signal of a tachogenerator; therefore it is potentially capable of contributing to extremely accurate digital speed control.

Encoders are available with an additional output known as a marker pulse, which identifies a specific orientation of the disc. By counting pulses from the datum provided by the marker pulse, the angle of rotation can be accurately measured. This feature is used in angular positioning systems, eg machine tool spindle orientation.

Absolute Encoder. The absolute encoder works in the same way as the incremental encoder described above, except that multiple channels are used to give a discrete code for every position increment. By decoding the output the angular position can be deduced directly, without the need to count from a datum or marker.

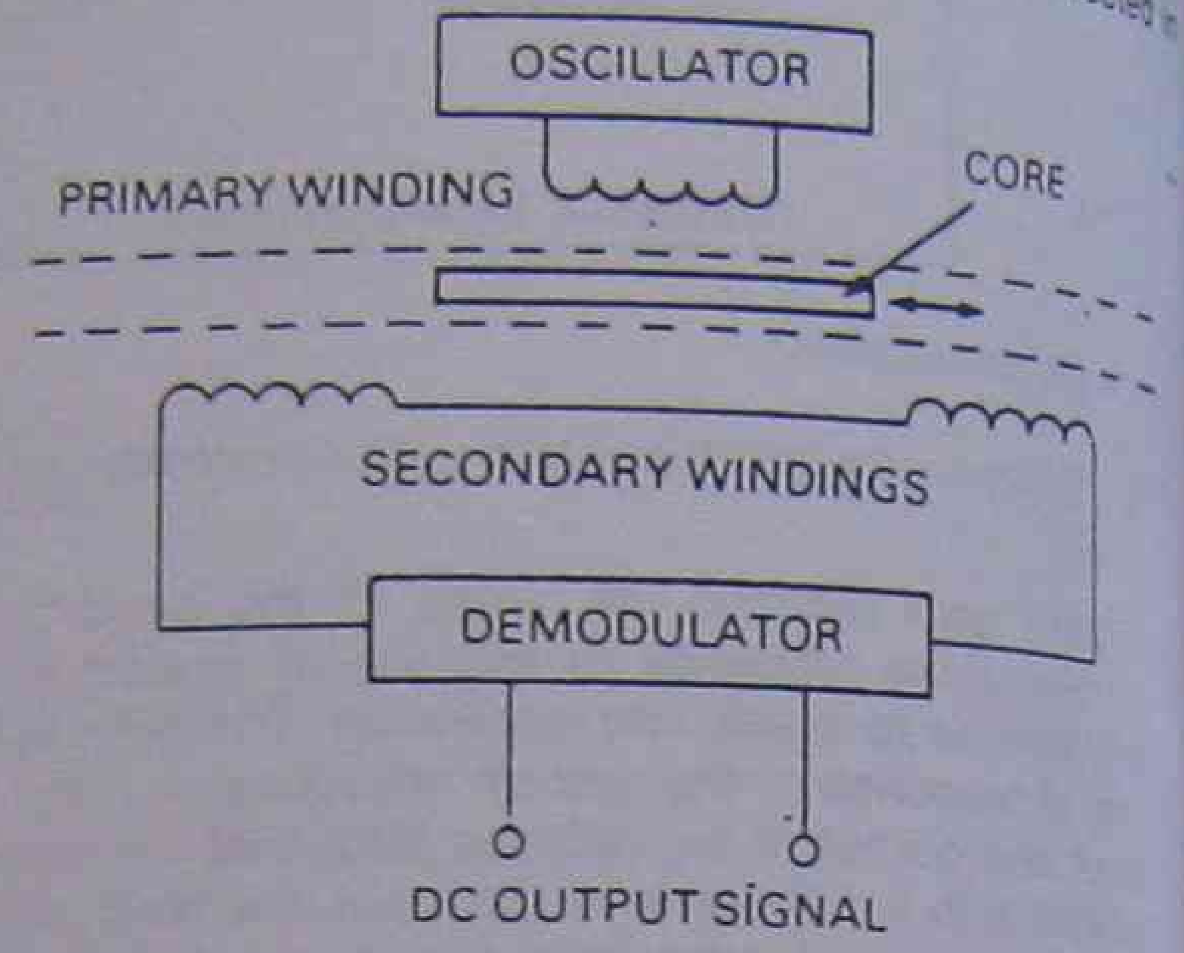
Absolute encoders are used in precision positioning systems, and are encountered more usually in conjunction with servo drives than with variable speed drives.

For the determination of linear position, linear encoders are available in both incremental and absolute configuration, having a grating in the form of a strip marked with transverse lines, which is free to move on its long axis relative to an optical sensor. They are used in precision and measurement positioning.

Resolver. The resolver is essentially a rotating transformer having a single primary winding, carried on the rotor, and two secondary windings in the stator, arranged at right angles to each other. The rotor position can be deduced from the relative magnitude of the secondary voltages and their phase relationship to the primary.

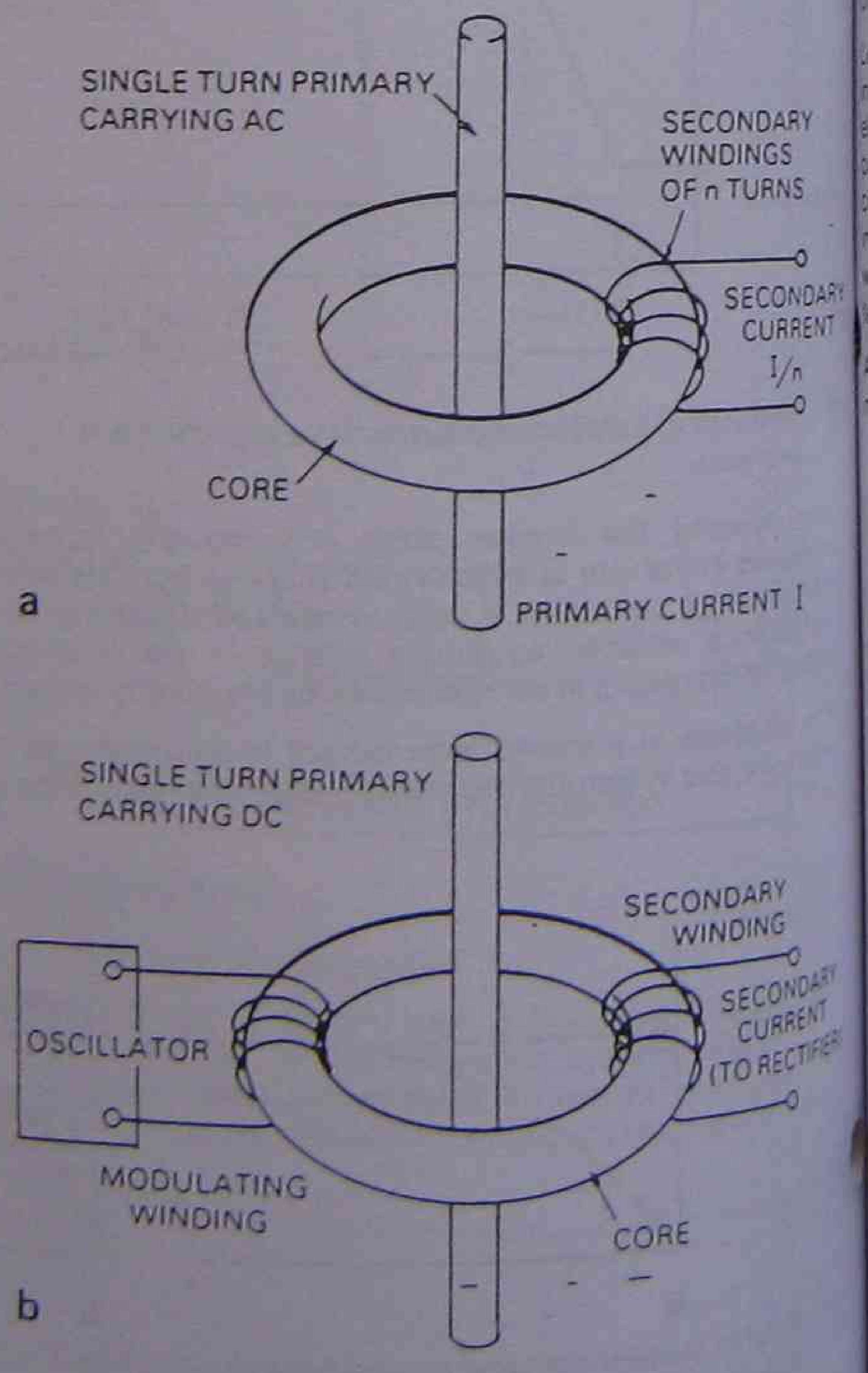
Linear Potentiometer. This is simply a resistive track with a sliding contact. The linear potentiometer is used in positioning applications which do not require the consistency, accuracy or long mechanical life of the more costly linear encoder.

Linear Variable Differential Transformer (LVDT). The LVDT is a linear position transducer, having three windings arranged side by side on a tubular former. The core is a cylindrical slug of iron or ferrite which moves within the former, as shown in Fig.7. The secondary windings are connected in opposition to one another, such that when the core is in the central position, the primary flux is coupled equally into both secondaries, and the output is zero. If the core is moved towards one end of the LVDT, the coupling is unequal and the secondary voltages have a resultant which is proportional to the distance the core has moved from its central (null) position.



7 *Linear variable differential transformer (LVDT)*

Shunt. A shunt is a low resistance connected in series with a current-carrying circuit. In accordance with Ohm's law, a voltage is developed across the shunt whose magnitude is proportional to the current flowing, and whose polarity depends on the direction of current flow. This voltage can be used as a feedback signal in current control loops.

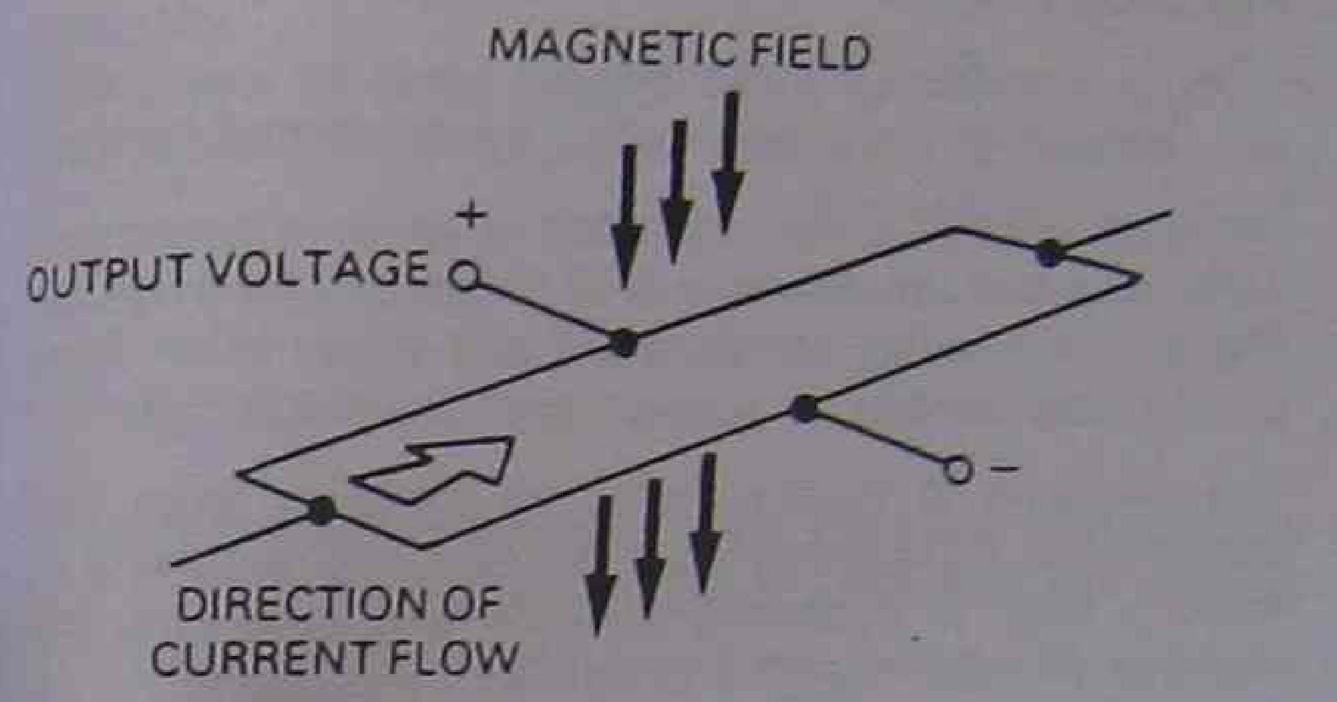


8 *Current transformers* **a** *(AC)CT* **b** *DCCT*

Current Transformers (CT and DCCT). A CT is usually a toroidal transformer, used for measuring alternating current. The output is a current signal whose magnitude is equal to the primary current divided by the turns ratio, Fig. 8a.

The DC Current Transformer (DCCT) incorporates an additional winding, connected to an oscillator, which modulates the flux in the core produced by the primary, in which direct current flows, Fig.8b.

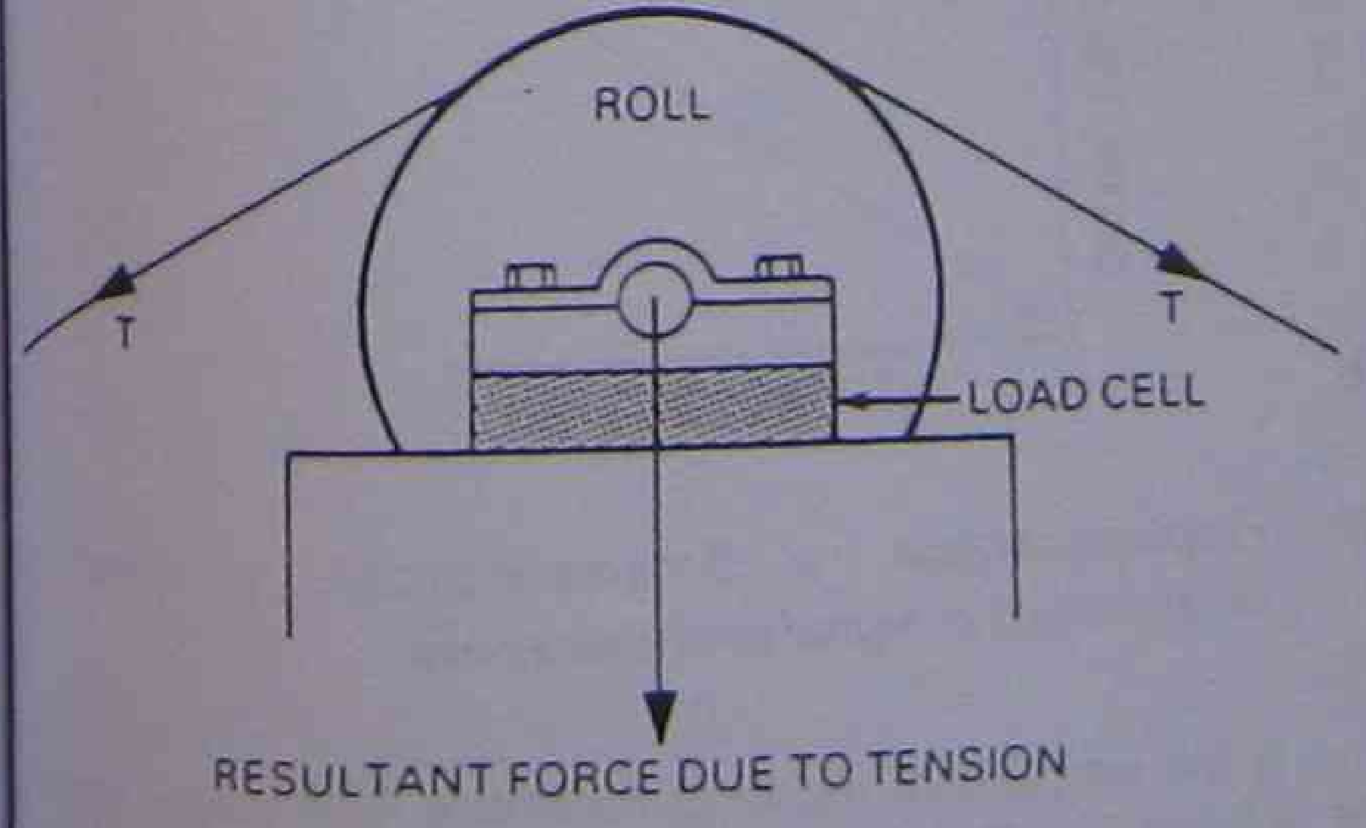
Hall-Effect Current Transducer. The Hall Effect, in which a voltage is developed across a current-carrying conductor placed in a magnetic field, is illustrated in Fig.9. If a Hall Effect transducer, fed by a constant-current source, is placed in a gap in a magnetic core having a current-carrying winding, an output voltage is produced which is proportional to the flux in the core and therefore to the current in the winding (provided that the core is operating in the linear part of its magnetisation characteristic).



9 *The Hall effect*

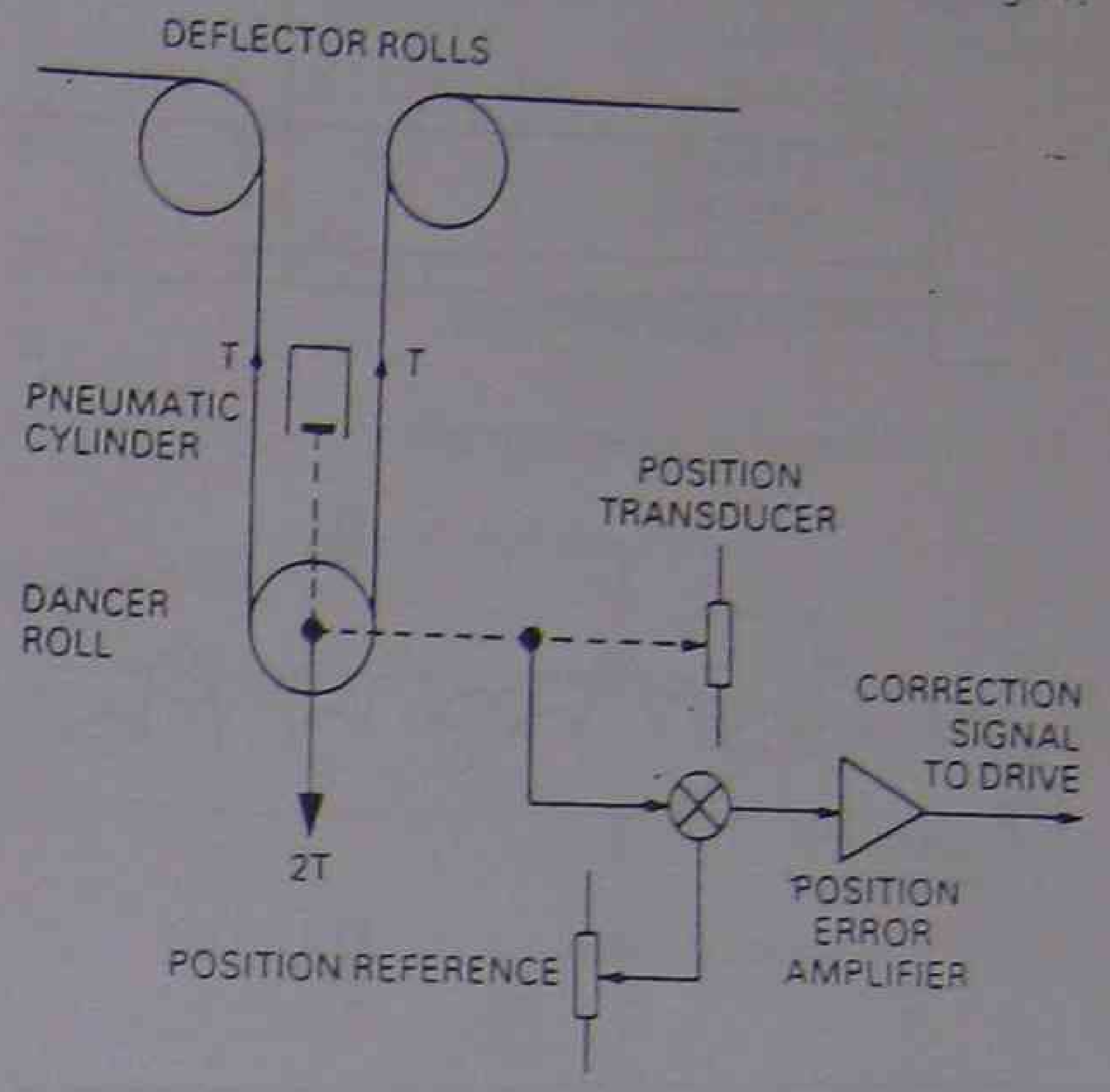
Load Cell. There are many types of load cell, some incorporating strain gauges, others incorporating piezoelectric transducers or LVDTs, but the purpose of each is to convert a mechanical force into an electrical signal. Load cells are often used in weighing, especially of bulk materials in hoppers, but the most important application in conjunction with thyristor drives is in the measurement and control of tension in continuous process lines.

A typical arrangement is shown in Fig.10, in which the material passes over a roll supported on load cells (often incorporated into the bearing pillow blocks; two are used, connected in series to compensate for uneven tension distribution across the material). Simple trigonometry is used to calculate the magnitude and direction of the resultant force on the load cell due to tension, and in most cases it is necessary to compensate for the tare weight of the roll and its bearings.



10 *Tension measurement by load cell*

Dancer. The dancer is itself not a feedback transducer, but is worth mentioning here because it is an example of indirect control — using position control as a means of controlling tension. A typical dancer arrangement is shown in Fig.11.

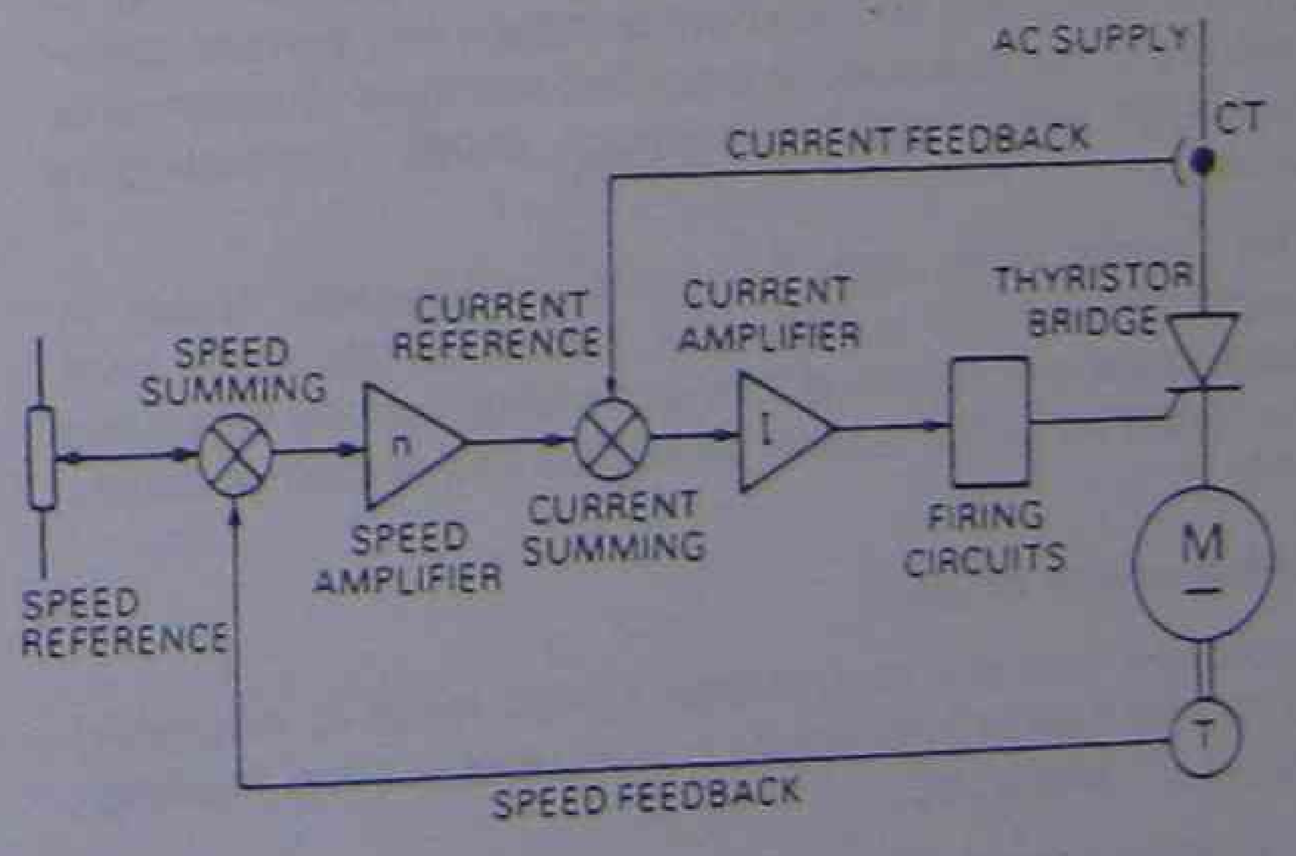


11 *Dancer tension control*

the material passing over a deflector roll, around the dancer roll which is free to move vertically, and returning to the original passline via a second deflector roll. The dancer is loaded (usually by a pneumatic cylinder) with a force which determines the tension in the material. Attached to the dancer is a position feedback transducer (often a linear potentiometer) and the position of the dancer is controlled via the line drives by means of a position loop. If the tension alters, the dancer rises or falls, causing a position error which is used to restore tension, and therefore the position of the dancer.

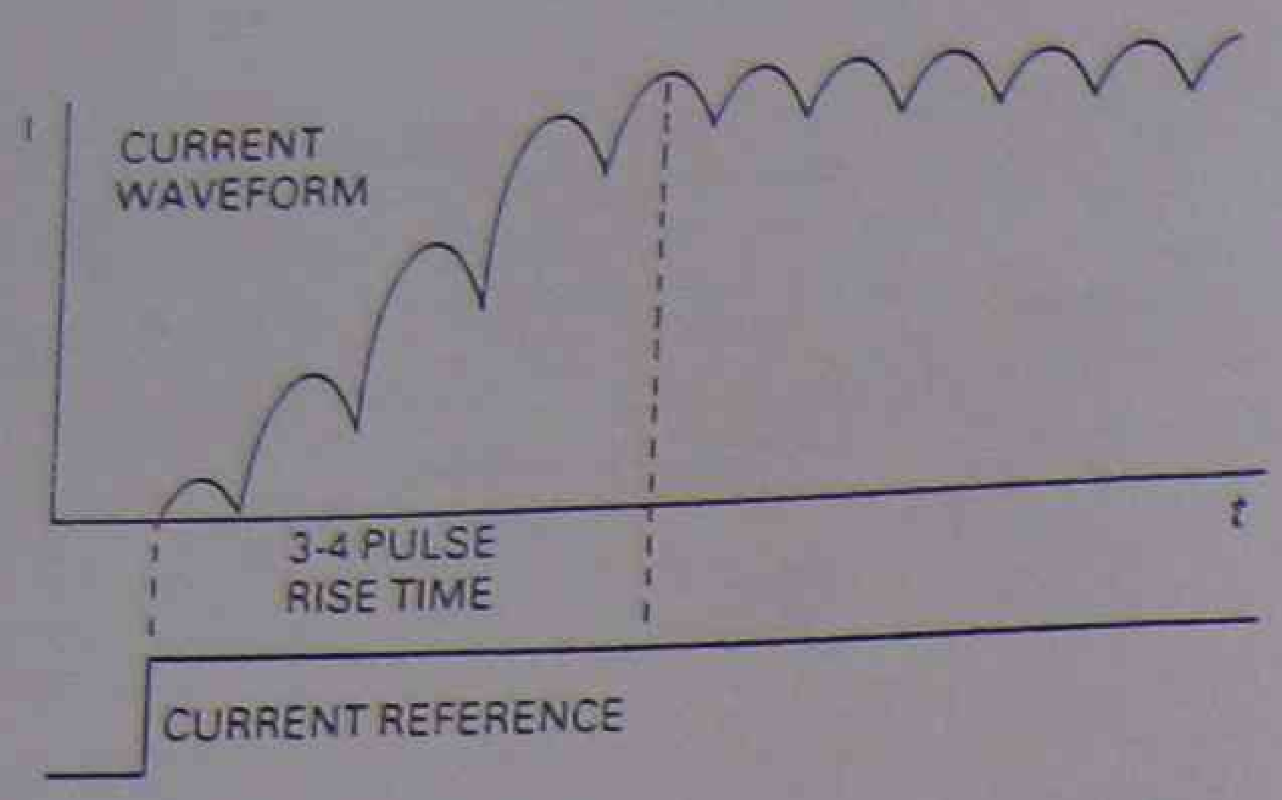
Cascaded Closed-Loop Control

Whilst the function of a variable speed drive in a closed-loop control system can be considered as the block labelled 'CONTROLLER' in Figs. 1 and 2, this is of necessity an oversimplification, and there are often two or more closed loops within the control system. For example, a drive in speed control, as illustrated by Fig.12, has not only a speed loop but also an inner current loop.

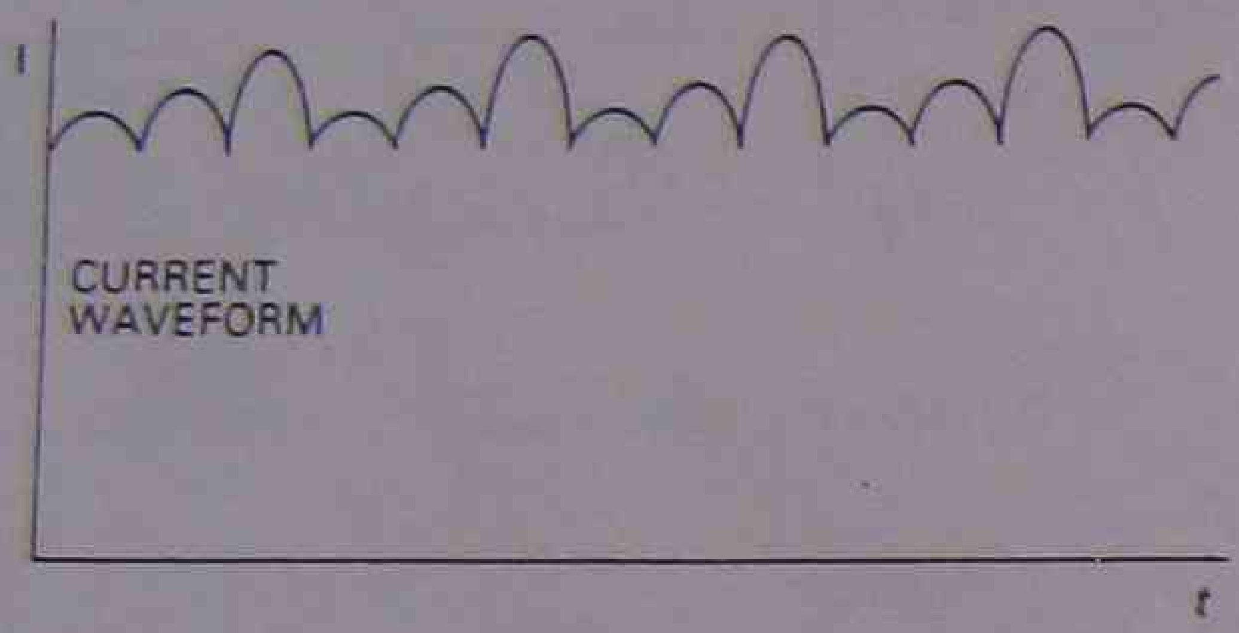


12 *Cascaded closed loop control of a thyristor drive*

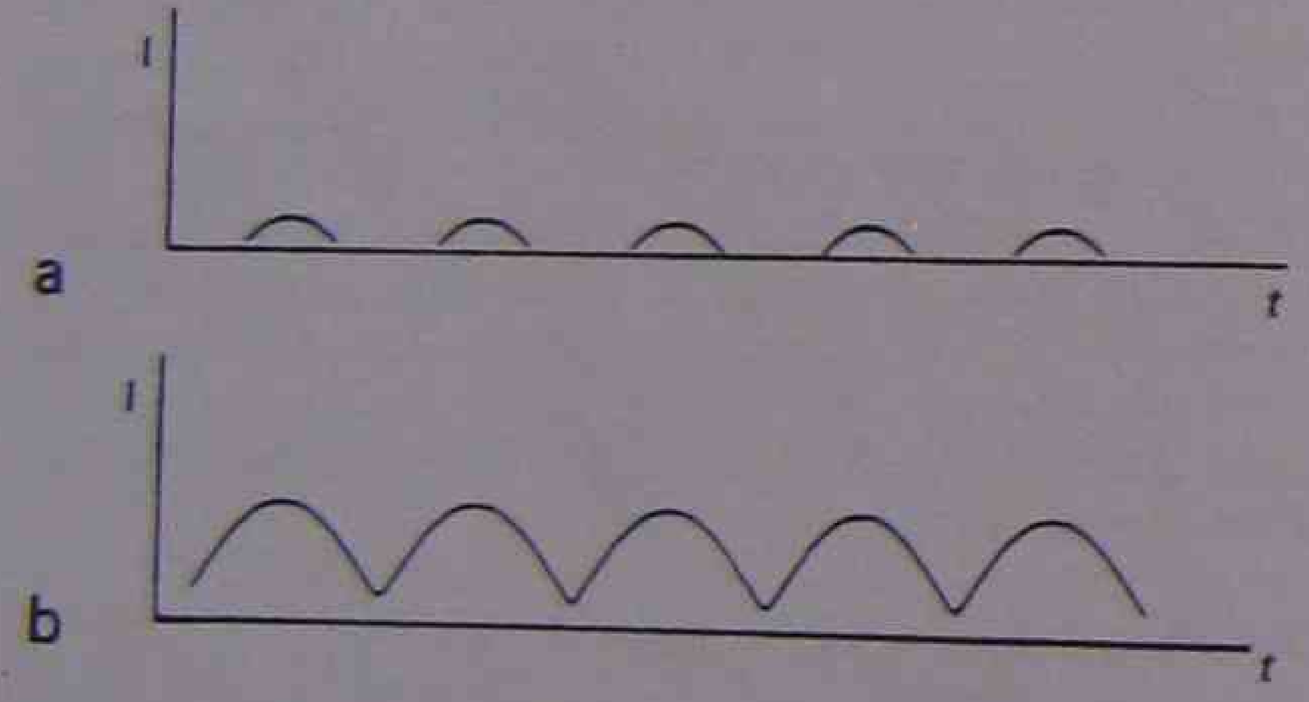
The speed error signal, resulting from the summing of reference and feedback, is amplified by the speed amplifier, and becomes the reference to the current loop. It is normal for limits to be imposed upon the magnitude of the current



13 *Optimisation of current loop gain*



14 *Current loop gain too high*



15 *Current waveforms* **a** *Discontinuous* **b** *Continuous current*

reference to prevent excessive current being delivered to the motor.

The current error signal, resulting from the summing of current reference with current feedback (via current transformers in this example), is amplified and used to advance or retard the firing angle of the bridge, thereby controlling its output.

The current loop can be made considerably faster in response than the speed loop, (which is limited by both mechanical and electrical time constants) thereby improving the response of the system to fluctuations in load or supply voltage.

Optimisation of gains

Excessive gain in a closed-loop system leads to instability, whilst insufficient gain results in poor response and 'hunting', as already explained.

In a drive having cascaded speed and current feedback loops, the inner current loop is the more critical and must be optimised before the speed loop is adjusted if the best performance is to be obtained. The actual gain required in the current loop depends on the electrical time-constant (L/R) of the motor armature circuit, and therefore varies from motor to motor. Customarily, the gain is adjusted by disconnecting the motor field supply and mechanically locking the shaft to prevent rotation, then applying a step input to the current loop whilst observing the current waveform by means of an oscilloscope. The current should rise to its final value after three to five pulses, Fig.13, although if an uneven waveform is seen, as in Fig.14, the gain is too high and should be reduced somewhat. The procedure is easier when dealing with modern, digital drives, since the gain is a precise digital quantity which can be calculated once the value of current reference required for continuous conduction has been ascertained. This is usually achieved by adjusting current limit with the motor stalled as above, until the desired waveform is obtained, Fig.15.

PID Loop Control

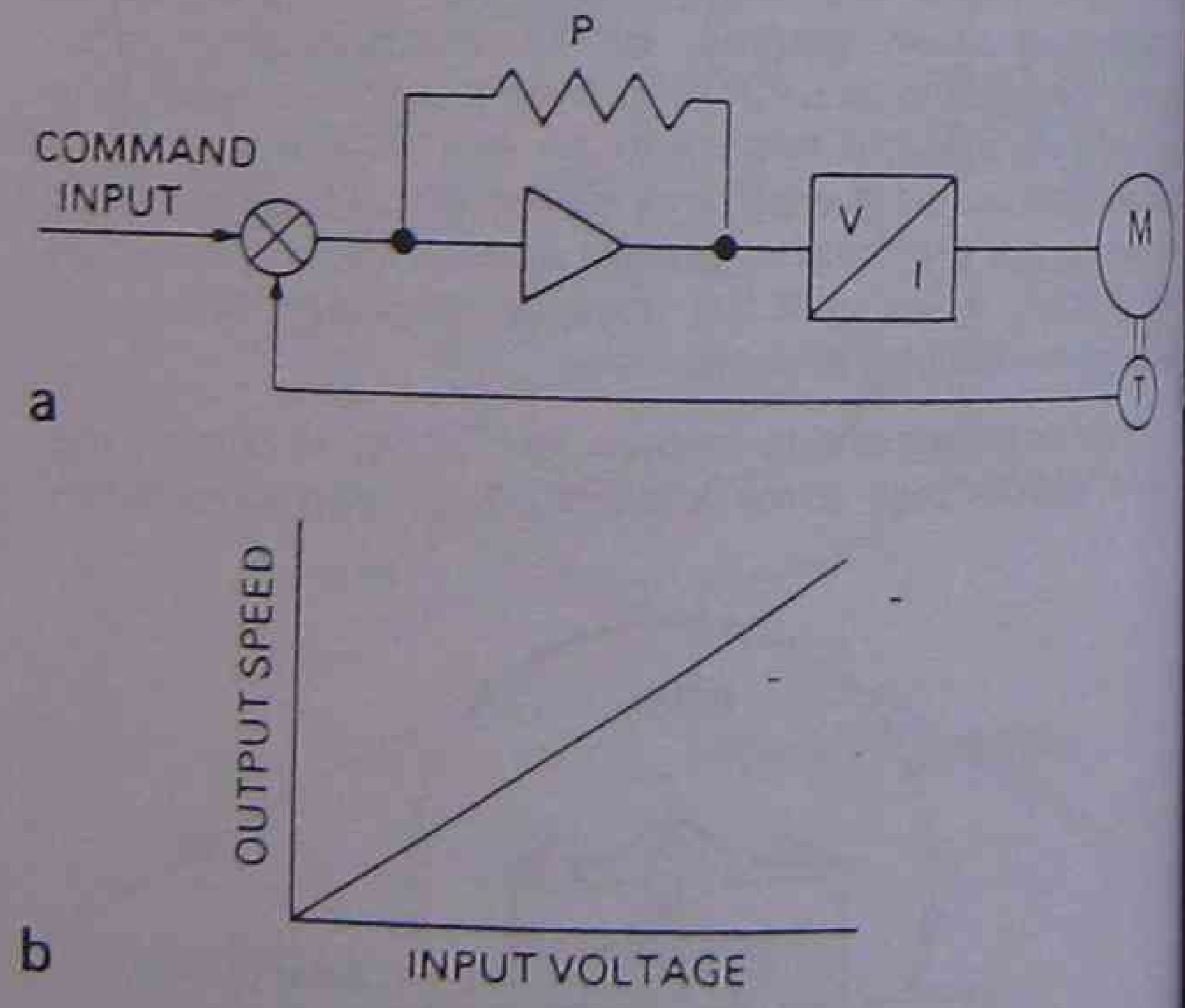
Two, or sometimes three distinct types of gain are to be encountered in the speed loop of variable speed drives.

PID loop control is not restricted to variable speed drive or servodrive control. Any one of all of the functions can also be found in hydraulic, temperature and many other forms of closed loop control.

The principle is illustrated by an example from variable speed drive technology to show how PID functions in the velocity loop of a servodrive, and how it can be used to improve and optimise the control of the speed of the motor. PID is a series of signal gains, each making a different adjustment to the feedback.

Proportional

Proportional gain is the factor by which the speed error is multiplied to produce the proportional speed correction term. Increasing proportional gain gives faster transient response and increased system damping, but leads to instability if carried to excess.



16 *Proportional gain:* **a** *Schematic circuit;* **b** *Linear relationships of signal to output speed.*

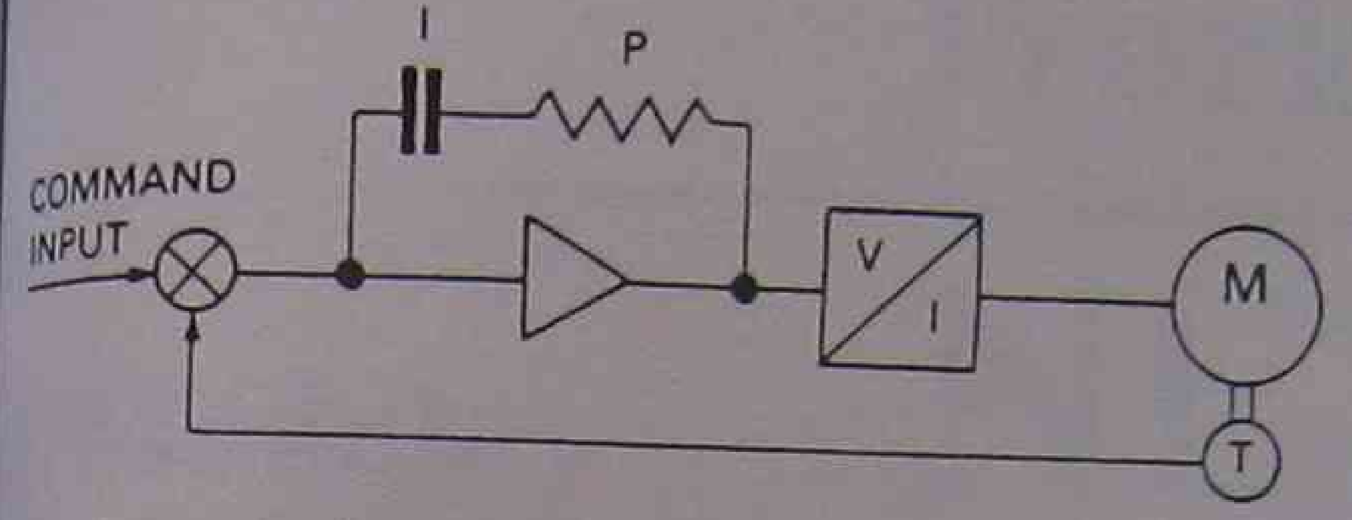
This is the simplest of the three functions. The output from the system, ie the speed, is made proportional to the external command signal input into the system. The electrical schematic for this function is shown in Fig.16a.

The relationship between the input signal and the output signal is linear, as shown in Fig.16b.

Integral

Integral gain is necessary to eliminate the speed error which would otherwise exist due to the fact that the proportional gain is finite. By integrating the speed error with respect to time, a correction term is produced which compensates for the initial speed error during steady-state operation. Increasing integral gain results in more rapid recovery after a transient disturbance (eg sudden load change) but again, instability is the result of excessive gain.

This function is incorporated to give the system good position-stiffness. In other words, when the drive is at stand-still, the integral function provides the control necessary to prevent rotation being induced by outside loads. This is achieved by incorporating a capacitor in the feedback of the velocity loop amplifier, Fig.17.



17 *Integral gain schematic circuit.*

At low speeds the reactance of the integrator capacitor is high. It therefore has more effect at zero and low speeds than at high speeds. i.e. almost open loop gain of the OP-AMP.

Derivative

Derivative gain, also known as forcing, is sometimes encountered. The error signal is differentiated to obtain its rate of change, which is then multiplied by the derivative gain to give a term which corrects for sudden transient disturbances. However, modern drives respond so rapidly that derivative gain is seldom necessary, and in addition, it has the undesirable effect of amplifying noise and ripple in the feedback signal, often contributing to instability.

The derivative function of the feedback loop helps to prevent overshoot when reaching the required speed, including zero speed. This is achieved by introducing as kind of feed-forward, which "predicts" the approach of the desired speed. (velocity feedback.)

Profile D, Fig.18a, is the sum of the derivative induced current plus the tacho feedback induced current.

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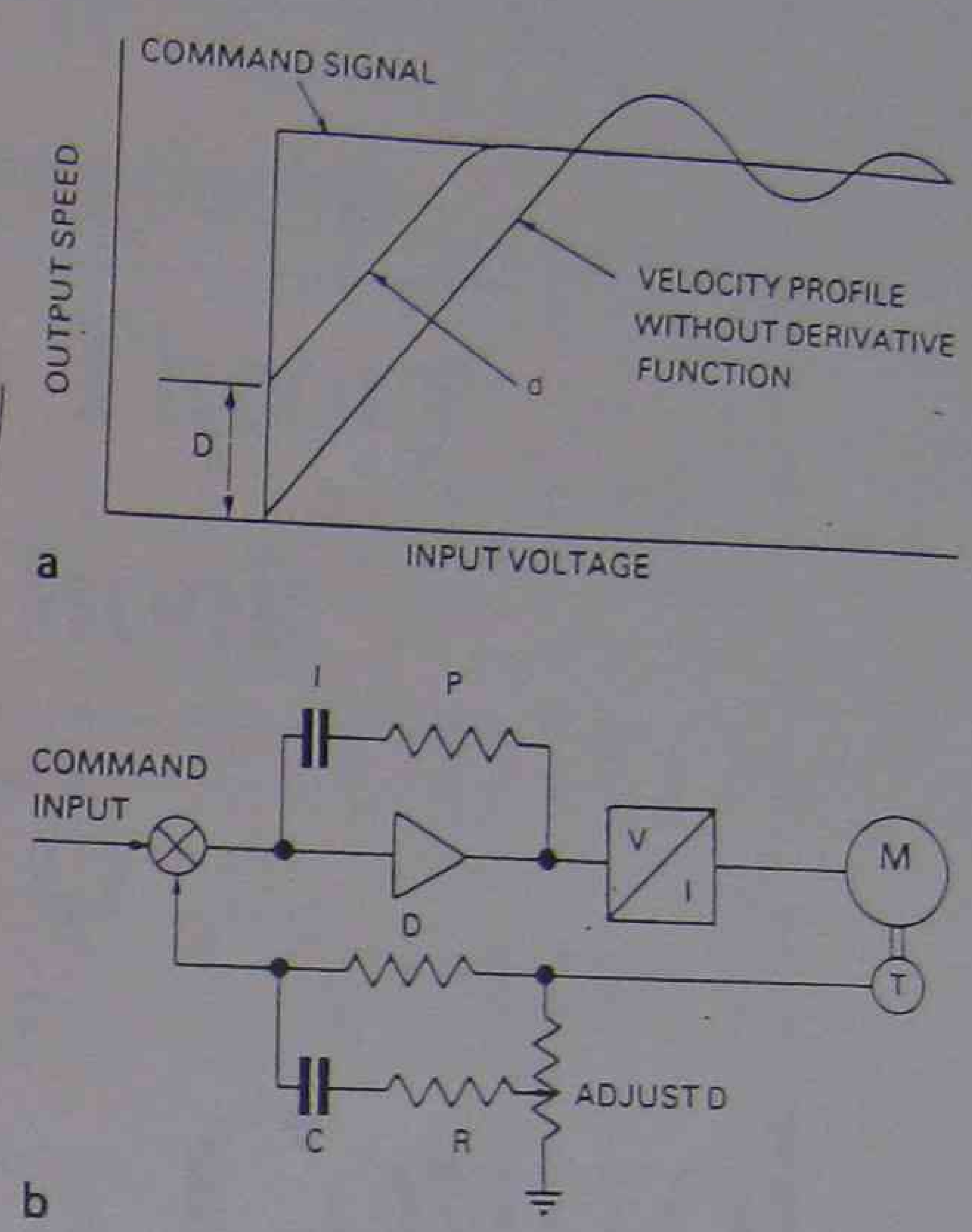

18 *Derivative gain:* **a** *Effect on control stability;* **b** *Schematic circuit.*

The electrical schematic for this function is shown in Fig.18b. In practice, the derivative function is adjustable so that the system can be optimized and incorporates the filter circuit as shown.

A simpler way to understand the derivative function is to imagine the temperature control of an electric oven. If the control of the oven were left until the desired temperature had been reached, then the temperature would rise above the set point. If however a derivative offset were introduced, then the control can be turned off before the set point is reached and the temperature will rise only to the required level.

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1

Principles of vector control

The first part of a two-part article which discusses the steady-state and dynamic operating characteristics of vector-controlled induction motors compared with dc motors for variable speed applications.

By Dr Dal Y Ohm

The induction motor is the most widely used electrical machine in industrial applications, mainly for constant-speed operations directly from the powerline. Induction motors have many advantages over dc motors, such as a simple and rugged structure, low cost, and reliability. However, for variable speed applications, induction motors have only recently found favour. Instead, dc motors have dominated because of relatively simple dc voltage control. For example, permanent magnet (PM) dc motors have been widely used for servo drives, and separately excited dc motors have been used in wide-speed-range applications, like machine tool spindles.

The popularity of dc motors in variable speed applications is not because the motor itself is inherently superior to ac motors. It is simply because the voltage and current of an ac power source is difficult to control, and proper control methods had not been developed.

With proper ac power control, ie, vector control, the induction motor can be as good as, or better than dc motors in high-performance, variable speed applications. For example, the ac motor has higher peak torque capability. The rapid evolution of power semiconductor and microprocessor technology has made induction motor with proper drives economically feasible. They can be applied in many applications, including machine-tool spindles, steel mills, etc. They can now challenge servo drives and traction motors.

Usually, high-performance drives use feedback control for fast dynamic response and to obtain accurate position or velocity. The motor and drive set in a motion control system serves as a torque amplifier. High-performance motion control can only be achieved if the torque characteristics of the motor and drive are stable and linear. Before discussing vector control, we'll compare the operational characteristics of dc and induction motors.

Dc motor

Figure 1 is a schematic of a PM dc motor connected to a load. Upon injection of voltage, V, armature current, la, flows in the rotating armature coil through the commutator and brush assembly. Mutual reaction between the stator's permanent magnet flux (F) and the rotor's armature current, la produces torque. Here:

$$T(t) = Ke\,F\,la(t) \qquad (1)$$

Back EMF, Vb, is produced as a load to the electrical system as:

$$Vb = Ke\,F\,\omega \qquad (2)$$

Where:

Ke = Constant determined by the motor structure

ω = Angular speed of the motor

The relationship between V and la is:

$$V(t) = L\,dla(t)/dt + Rla(t) + \cancel{Vb(t)}\ Vb(t) \qquad (3)$$

Where:

L = Motor inductance

R = Armature resistance

From Equation (1), torque is proportional to armature current if flux stays constant. When supply voltage, V, is controlled, la has a time delay of L/R, and its magnitude is dependent on speed (due to back EMF term). If la is directly controlled, these problems can be eliminated. Therefore, most high-performance drives use a current amplifier that is either a current source or PWM amplifier with high gain current feedback.

From Equation (2), Vb is proportional to motor speed. If this voltage approaches the supply voltage, current (and torque) can no longer be controlled because of the small potential difference available to the amplifier. Usually this is the operational speed limit of a PM dc drive; it is called "base speed".

Next, consider a separately excited dc motor whose flux is supplied from field current connected from the independent power source, instead of a permanent magnet. Here, flux is proportional to field current, lf. Equation (1) can then be expressed, with a different constant:

$$T(t) = Kd\,lf\,la(t) \qquad (4)$$

When field current is fixed at the highest designed value, its operational characteristics are identical with that of a PM motor, and speed is limited by supply voltage. But, if field current is allowed to change, higher speed operation is possible with a reduction in field current. For example, if field current (and flux F)

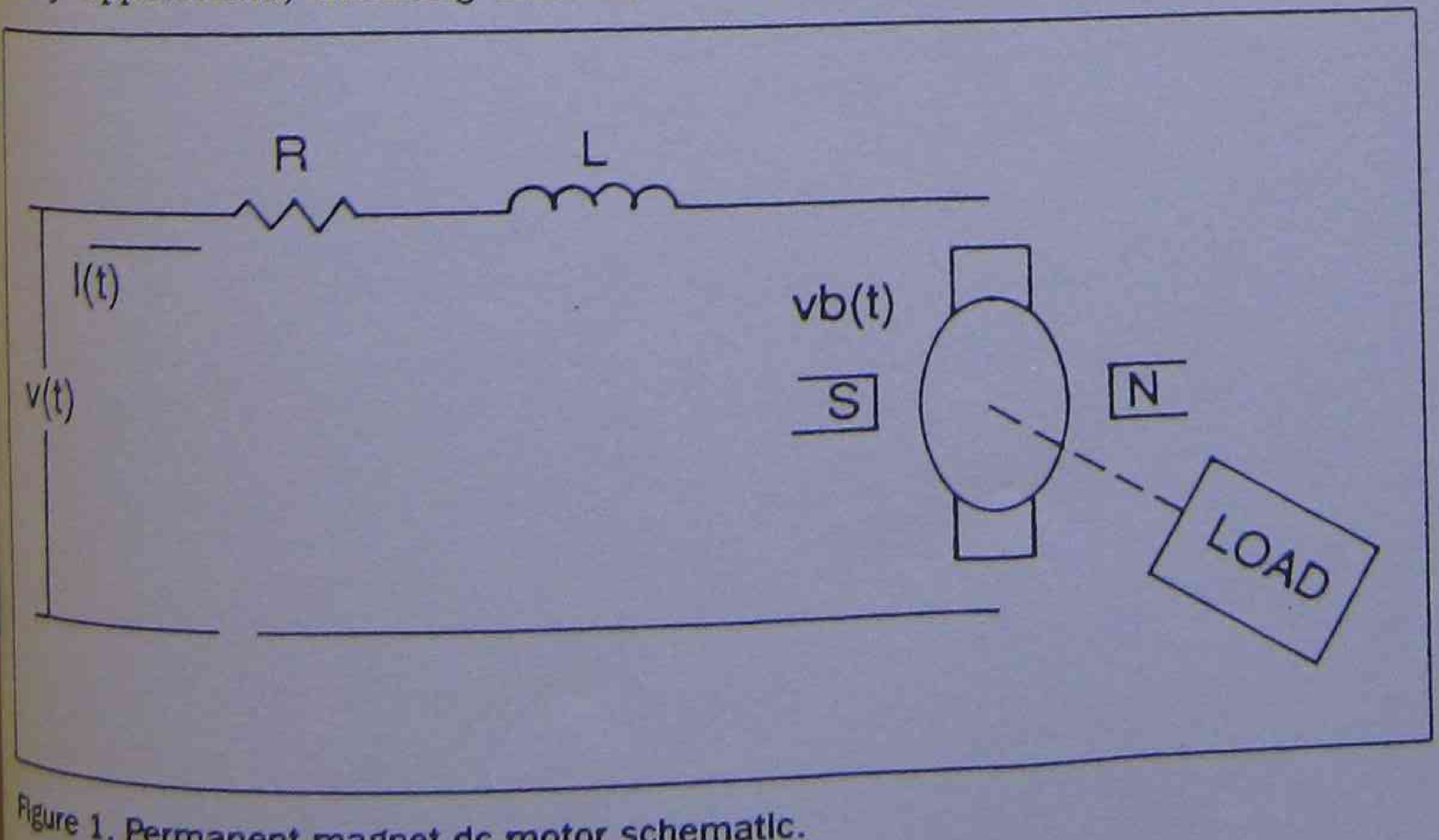

Figure 1. Permanent magnet dc motor schematic.

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are cut in half, back EMF is also reduced by half (Equation (2)). Therefore, the motor can run up to twice the previous speed because of the additional voltage available for armature current control. Here, torque constant, Ke, is reduced in half.

One popular method is to control field current inversely proportional to the speed above base speed. Operational speed range above base speed is called the "constant power region" because available power (T(t)* ω(t)) is constant in this region. Figure 2 shows torque and power characteristics of a separately excited dc motor. Theoretically, speed can be increased indefinitely, but in practice speed is usually limited by the commutator and brush design, system frictions etc. A special motor must be designed for high-speed operation. Loads like machine tool spindles, rolling mills or traction applications require low torque at higher speed, and this field-weakening method has been widely used.

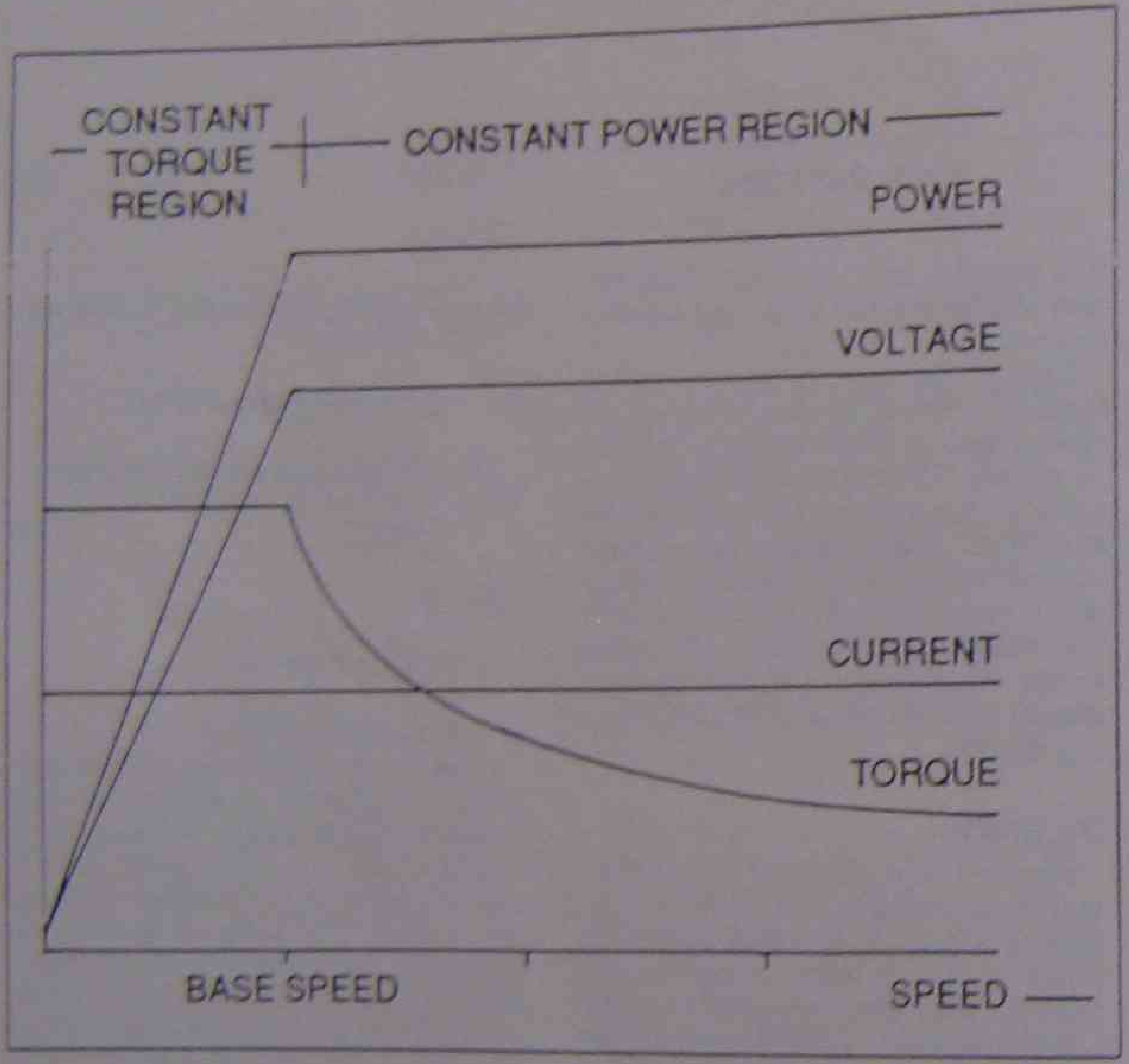


Figure 2. Separately excited dc motor characteristics.

Dc motors are still widely used in industry, but their commutators require periodic maintenance; they also have low reliability and are dangerous in explosive environments. Another disadvantage of the dc motor is that its commutator or demagnetisation limits its peak torque. Besides these drawbacks, high-power dc motors are much more expensive than induction motors.

Three-phase Induction motor

Three-phase induction motors have an ac winding on the stator so that spatially distributed magnetic flux rotating around the airgap is generated from the supplied ac voltage. For single-phase motors magnetic flux is not rotating but stationary and pulsating. Poor performance of single-phase induction motors, especially at low speed, precludes their use in most variable speed applications. From now on, only polyphase (three-phase in particular) motors will be discussed. If ac voltage (with constant frequency, f_o) is supplied to the stator winding of an induction motor, the stator generates rotating magnetic flux, revolving at a speed, *no*, (RPM) determined by:

$$no = \frac{60f_o}{P} \text{ [RPM]} \quad (5)$$

where:

P = Number of motor pole pairs

This speed is called synchronous speed, and induction motor speed at no load is very close to this speed. Equation (5) also is a general equation relating electrical frequency to mechanical speed.

3

In cage-type polyphase induction motors, the rotor coil is simply several conductor bars connected at both ends, and transformer action induces rotor current from the stator. Mutual reaction between the rotating magnetic flux and rotor current produces torque. Consider that the rotor is turning at a speed *nr*, which is less than *no* during normal motoring operation. Frequency fs (slip frequency) of the rotor circuit is determined by:

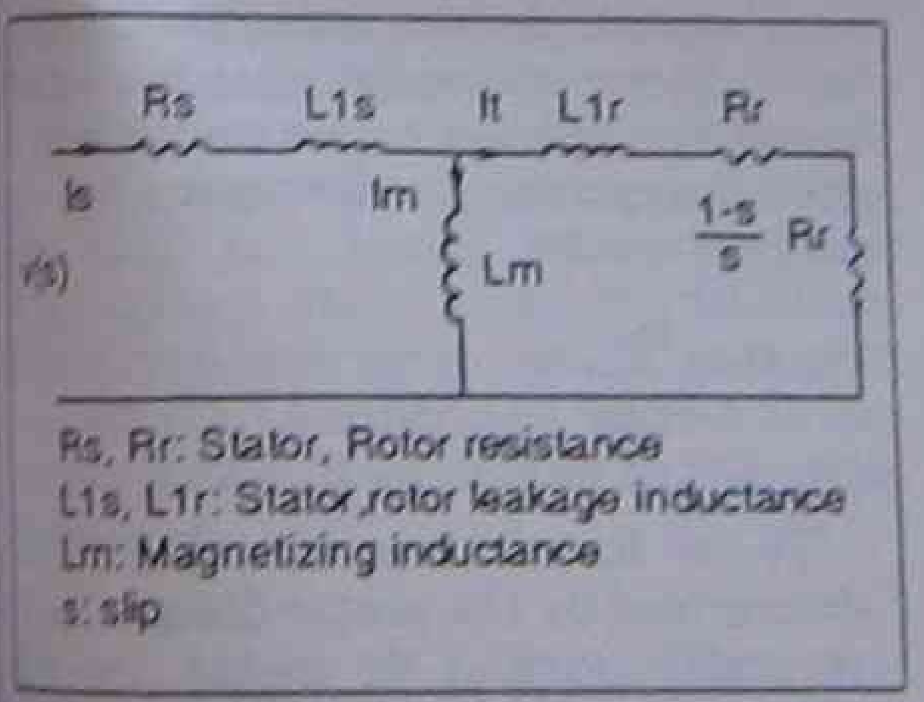
$$f_o = f_r + f_s \quad (6)$$

where:

f_r = Speed *nr* converted to frequency using Equation (5). We can define angular frequencies ω_o, ω_r and ω_s that correspond to f_o, f_r and f_s multiplied by 2π. Sometimes per-unit slip, s, may be used as defined by:

$$s = \frac{fs}{fo} = \frac{no - nr}{no} \quad (7)$$

Steady-state characteristics of an induction motor are best described by its per-phase equivalent circuit shown in Figure 3. In this equivalent circuit, rotor circuit parameters are referred to as stator side via tranformation. Produced torque can be expressed by:



Rs, Rr: Stator, Rotor resistance
L1s, L1r: Stator,rotor leakage inductance
Lm: Magnetizing inductance
s: slip

Figure 3. Steady-state equivalent circuit of induction motor.

$$T = Ka\, lm(t)\, lt(t) \sin \alpha \quad (8)$$

where:

α = Phase difference between lm(t) and 1q(t)

From Equation (7), maximum torque is produced when α is almost 90° (slip s is close to zero and the effect of leakage inductance is small). When the motor starts (s = 1), the phase difference between two current components is very small because leakage inductance is dominant in the rotor circuit. The angle is small, and large rotor current (inrush current) flows in the motor to produce the require torque.

Torque characteristics of an induction motor operating directly from the power line (constant voltage, constant frequency ac source) is shown as Curve A in Figure 4. Stable operation is between the synchronous speed to maximum torque point (marked as "M") and the rated operating point marked with an "R". As shown in the figure, the stable operating speed is about 90% to 100% of synchronous speed. If the power source frequency is allowed to change, this stable operating range can be moved up and down. Flux current lm(t) is roughly determined from:

$$lm(t) = \frac{V(t)}{(\omega_o Lm)} \quad (9)$$

Constant flux can be maintained if the ratio of supply voltage per frequency (known as V/H) is preserved, and the rated and maximum torque values stay the same. Curve B represents the torque characteristics when both voltage and frequency are reduced in half. "Inverter" is a variable frequency, variable voltage power supply that maintains the V/Hz ratio. If the supply frequency increases supply slowly to a desired value from a very low frequency, the induction motor always starts and runs at a stable operating speed. Here, inrush current does not flow because slip is always small. As in a separately excited dc motor, high-speed, constant-power operation is possible with increasing frequency while maintaining voltage at the rated voltage above base speed. Curve C and D in Figure 4 illustrates torque curve above base speed. Steady state torque characteristics of an inverter driven induction motor is very similar to Figure 2. With inverter operation, unstable operation might result if acceleration or deceleration is too fast, and motor speed cannot track the frequency within a stable slip value. Therefore, an inverter driven induction motor is used only in applications where fast dynamic performance is not important.

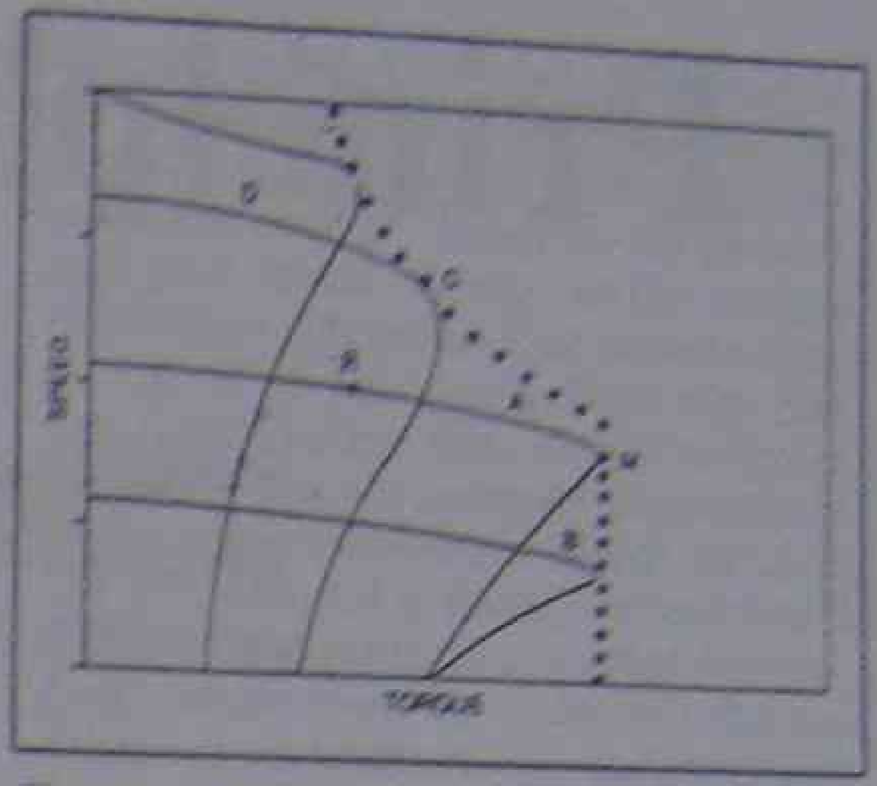


Figure 4. Torque-Speed curves of inverter-driven induction motor.

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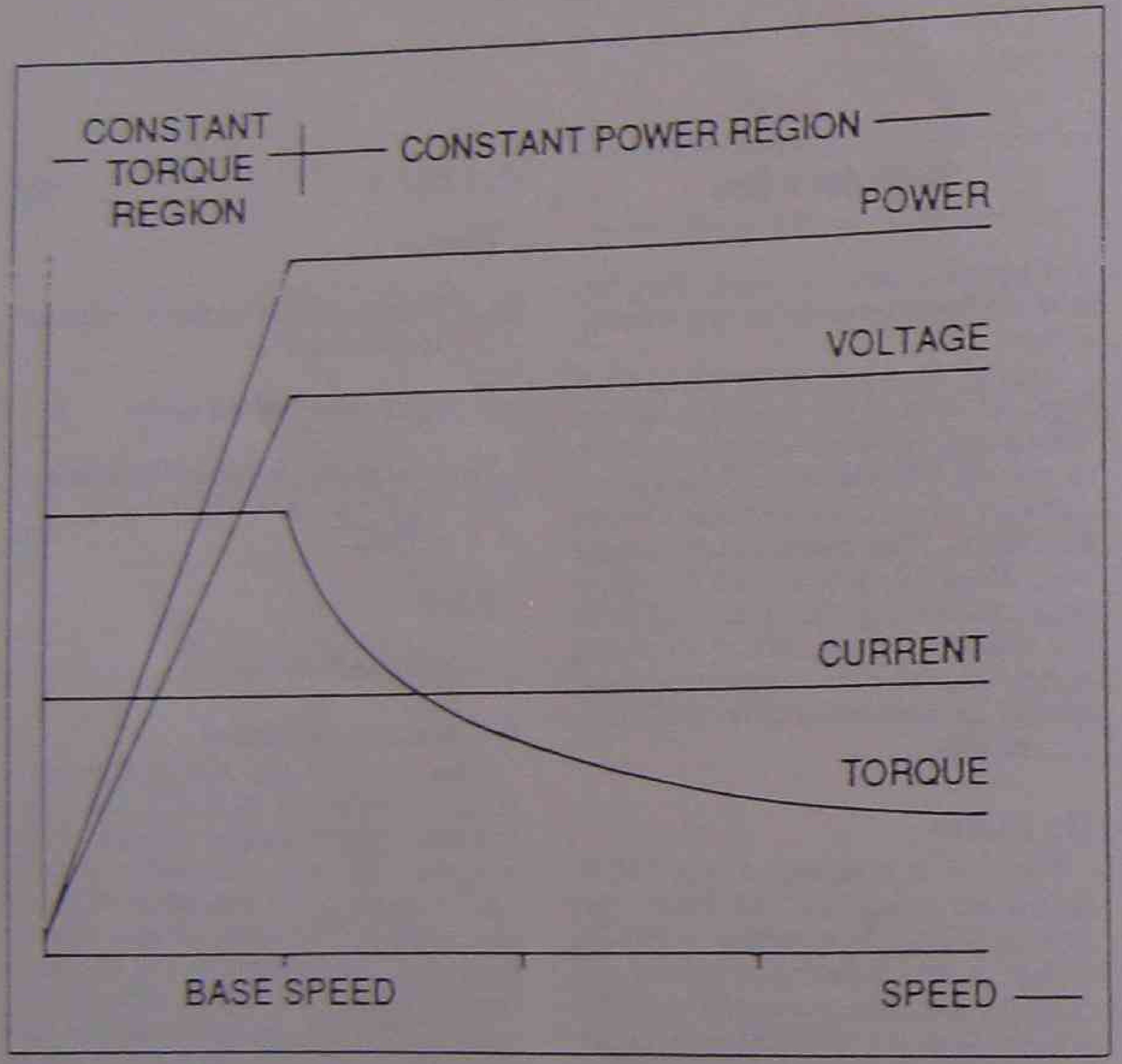

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Dc motors are still widely used in industry, but their commutators require periodic maintenance; they also have low reliability and are dangerous in explosive environments. Another disadvantage of the dc motor is that its commutator or demagnetisation limits its peak torque. Besides these drawbacks, high-power dc motors are much more expensive than induction motors.

Three-phase Induction motor

Three-phase induction motors have an ac winding on the stator so that spatially distributed magnetic flux rotating around the airgap is generated from the supplied ac voltage. For single-phase motors magnetic flux is not rotating but stationary and pulsating. Poor performance of single-phase induction motors especially at low speed, precludes their use in most variable speed applications. From now on, only polyphase (three-phase in particular) motors will be discussed. If ac voltage (with constant frequency, f_o) is supplied to the stator winding of an induction motor, the stator generates rotating magnetic flux, revolving at a speed, *no*, (RPM) determined by:

$$no = \frac{60 f_o}{P} \text{ [RPM]} \qquad (5)$$

where:

P = Number of motor pole pairs

This speed is called synchronous speed, and induction motor speed at no load is very close to this speed. Equation (5) also is a general equation relating electrical frequency to mechanical speed.

In cage-type polyphase induction motors, the rotor coil is simply several conductor bars connected at both ends, and transformer action induces rotor current from the stator. Mutual reaction between the rotating magnetic flux and rotor current produces torque. Consider that the rotor is turning at a speed *nr*, which is less than *no* during normal motoring operation. Frequency fs (slip frequency) of the rotor circuit is determined by:

$$f_o = f_r + f_s \qquad (6)$$

where:

f_r = Speed *nr* converted to frequency using Equation (5). We can define angular frequencies ω_o, ω_r and ω_s that correspond to f_o, f_r and f_s multiplied by 2π. Sometimes per-unit slip, s, may be used as defined by:

$$s = \frac{fs}{fo} = \frac{no - nr}{no} \qquad (7)$$

Steady-state characteristics of an induction motor are best described by its per-phase equivalent circuit shown in Figure 3. In this equivalent circuit, rotor circuit parameters are referred to as stator side via tranformation. Produced torque can be expressed by:

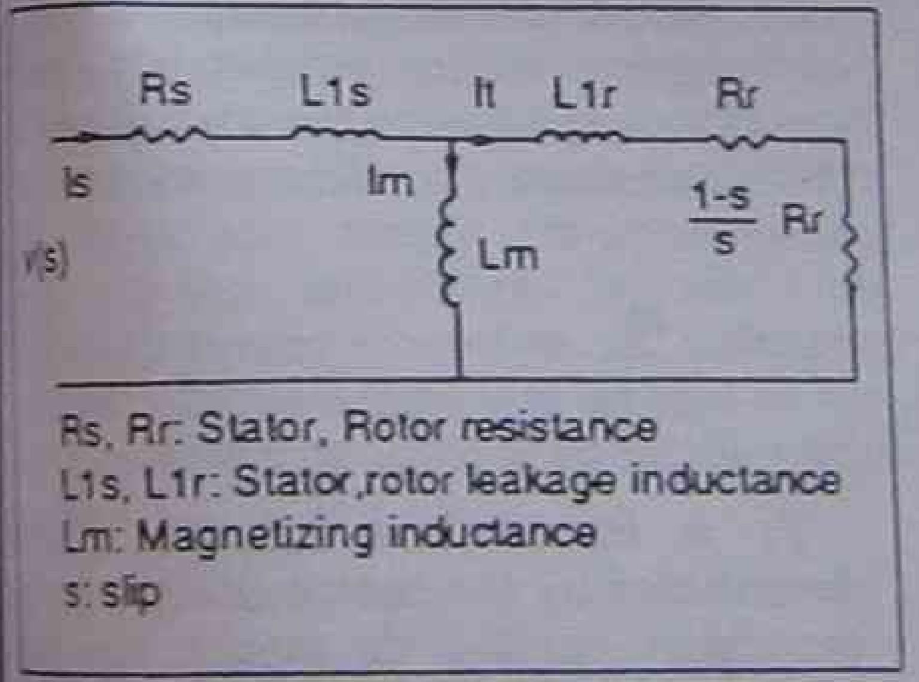

Figure 3. Steady-state equivalent circuit of induction motor.

$$T = Ka\, Im(t)\, It(t) \sin \alpha \qquad (8)$$

where:

α = Phase difference between Im(t) and 1q(t)

From Equation (7), maximum torque is produced when α is almost 90° (slip s is close to zero and the effect of leakage inductance is small). When the motor starts (s = 1), the phase difference between two current components is very small because leakage inductance is dominant in the rotor circuit. The angle is small, and large rotor current (inrush current) flows in the motor to produce the require torque.

Torque characteristics of an induction motor operating directly from the power line (constant voltage, constant frequency ac source) is shown as Curve A in Figure 4. Stable operation is between the synchronous speed to maximum torque point (marked as "M") and the rated operating point marked with an "R". As shown in the figure, the stable operating speed is about 90% to 100% of synchronous speed. If the power source frequency is allowed to change, this stable operating range can be moved up and down. Flux current Im(t) is roughly determined from:

$$Im(t) = \frac{V(t)}{(\omega_o Lm)} \qquad (9)$$

Constant flux can be maintained if the ratio of supply voltage per frequency (known as V/H) is preserved, and the rated and maximum torque values stay the same. Curve B represents the torque characteristics when both voltage and frequency are reduced in half. "Inverter" is a variable frequency, variable voltage power supply that maintains the V/Hz ratio. If the supply frequency increases supply slowly to a desired value from a very low frequency, the induction motor always starts and runs at a stable operating speed. Here, inrush current does not flow because slip is always small. As in a separately excited dc motor, high-speed, constant-power operation is possible with increasing frequency while maintaining voltage at the rated voltage above base speed. Curve C and D in Figure 4 illustrates torque curve above base speed. Steady state torque characteristics of an inverter driven induction motor is very similar to Figure 2. With inverter operation, unstable operation might result if acceleration or deceleration is too fast, and motor speed cannot track the frequency within a stable slip value. Therefore, an inverter driven induction motor is used only in applications where fast dynamic performance is not important.

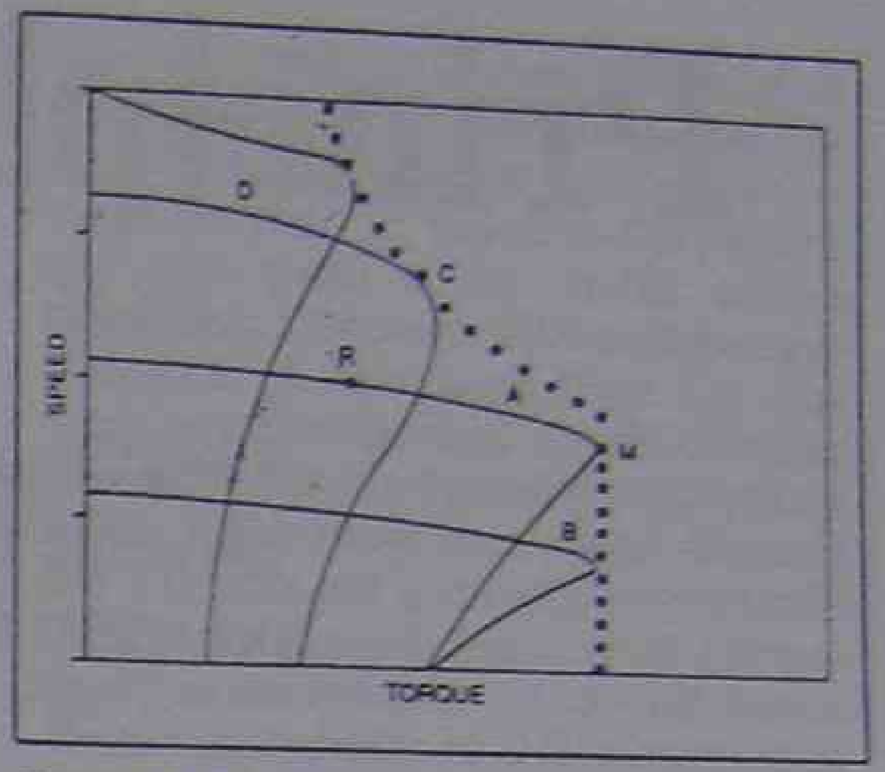

Figure 4. Torque-Speed curves of Inverter-driven Induction motor.

Principles of vector control

Part two of an article discussing steady-state and dynamic operating characteristics of vector controlled induction motors compared with dc motors for variable speed applications.

By Dr Dal Y Ohm

Vector control principles are based on the synthesis of ac induction motor current. Consider equations (4) and (8):

$$T(t) = Kd\,If\,Ia(t) \quad \text{(dc Motor)} \qquad (4)$$
$$T = Ka\,Im(t)\,It(t)\sin\alpha \quad \text{(Induction)} \qquad (8)$$

where:

α = Phase difference between Im(t) and Iq(t)

For two different types of motors, torque equations are very similar except the term sin α. In the dc motor, this angular relationship is still present but is fixed to 90° by the motor design. If the angular position of brushes is allowed to vary, corresponding sinusoidal terms can be applicable in a dc motor, too. The principle of vector control is to synthesise stator current so that these two component currents maintain orthogonality and are controlled independently in all dynamic situations. Decoupling these two current components maximises the efficiency of the produced torque, and provides operational characteristics similar to the dc motor.

The difference between the dc motor and the induction motor is that in the induction motor the airgap flux rotates whereas it is stationary in the dc motor. Because rotor current, or rotor mmf, produced from this current travels past the rotor at slip speed, and slip frequency EMF is generated in the rotor circuit by the transformer action, the relative angular position of rotor mmf is stationary relative to the stator mmf.

If we use a synchronously rotating reference frame as a reference coordinate system, both the stator and rotor fluxes are stationary and their mutual interaction generates continuous torque.

A mathematical linear transformation is possible between a three phase stationary reference frame and a two phase rotating reference frame with d and q axes. In the rotating reference frame, variables like voltages or currents are purely real quantities (like dc quantities) without modulation. In this synchronous frame, we can define Id and Iq, which correspond to rms values of Im(t) and It(t) if leakage inductances are neglected. Thus, equation (8) can be written as:

$$T = K\,Id\,Iq\sin\alpha \qquad (10)$$

Here, an analogy to the dc motor is possible, where Id corresponds to field current, and Iq corresponds to armature current. Because the two current components are orthogonal to each other (Id lags Iq by 90°), stator current is the vector sum, ie:

$$I_s = \sqrt{Id^2 + Iq^2} \qquad (11)$$

These vector relationships are shown in Figure 5. In the induction motor, Iq and slip are mutually dependent quantities. Once Iq is set by the torque requirement, angular slip frequency ω_s must satisfy:

$$\omega_s = \frac{Iq}{Tr\,Id} \qquad (12)$$

where

Tr = Rotor time constant (Lr/Rr) determined by the motor design. As in a separately excited dc motor, the magnitude of magnetising current, Id, is fixed to a constant value below base speed and reduced inversely proportional to speed above base speed. Torque producing current Iq is controlled to meet the torque requirement of the load.

To supply synthesised stator current to the motor, Id and Iq must be transformed back to three phase sinusoidal phase currents. This inverse transformation requires information on the instantaneous angular position of the flux. Two methods can be considered in obtaining this airgap flux position. "Direct vector control" actually measures by using flux sensors. The "indirect method" estimates it by integrating rotor speed and generated slip. The indirect method is more popular because measurement of flux requires additional hardware and complex signal processing. From now on, discussions will be limited to the indirect method. Here, the instantaneous phase of stator current is calculated by:

$$\varphi = \int (\omega_r + \omega_s)\,dt + \theta \qquad (13)$$

where:

$$\theta = \text{Atan}\,(Iqs/Ids) \qquad (14)$$

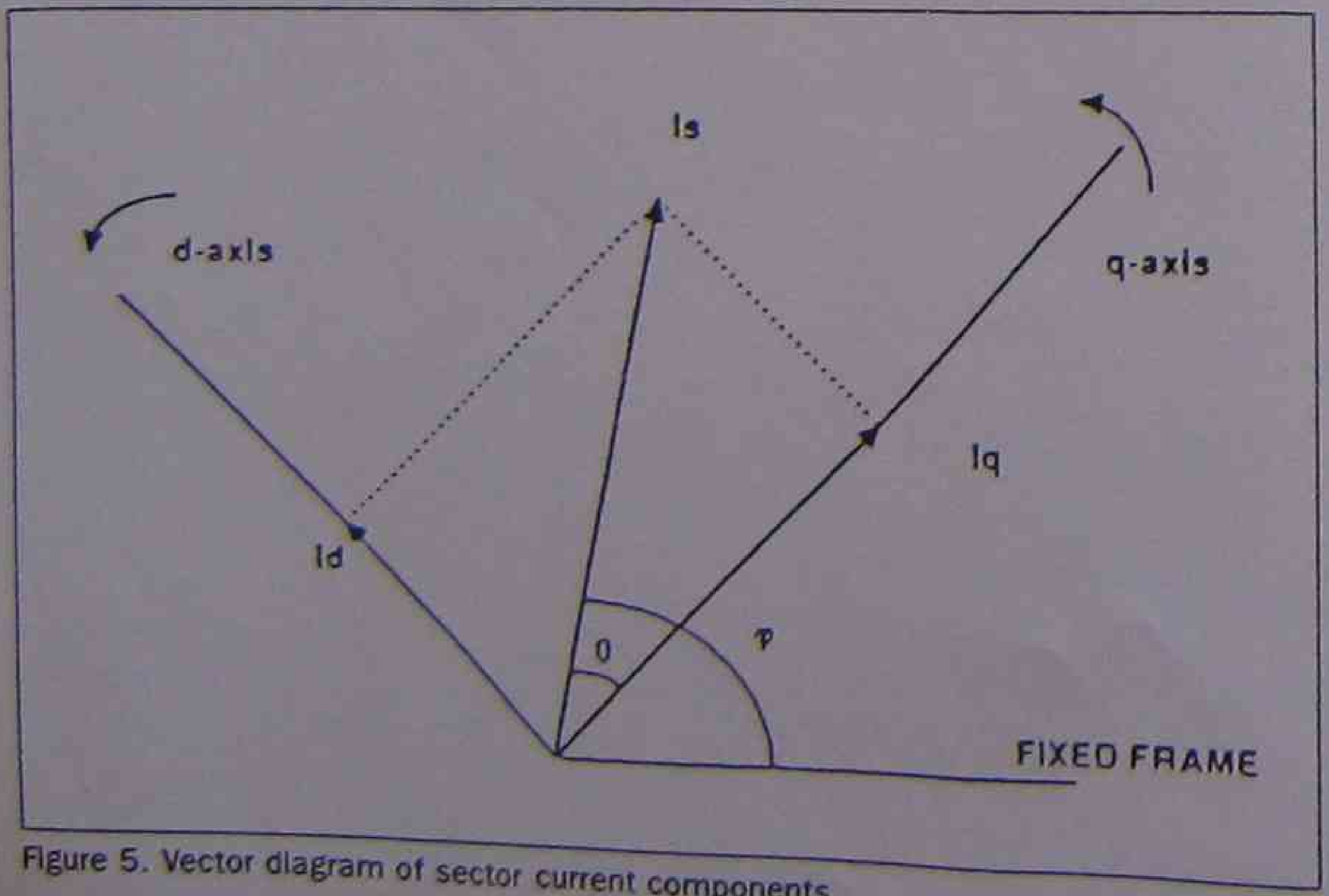

Figure 5. Vector diagram of sector current components

Note that stator current phase current is dependent upon slip, which is based on the knowledge of Tr in Equation (12). If the estimated Tr is not matched to the actual Tr, the two current components are not orthogonal. Therefore, the produced torque will be less than expected, resulting in poor efficiency and the inability to produce maximum torque. So, in vector control, correct tuning of parameter Tr is very important. With correctly tuned Tr, produced torque is:

$$T = Kv\,Id\,Iq \qquad (15)$$

where:

Kv = (3/2)P(Lm/Lr). Clearly, the above equation is analogous to the dc machine equation (equation (4)). Torque characteristic curves are depicted in Figure 6. Note that the stator current is slightly reduced as speed increases above base speed. This is due to the vector relationship in equation (11). The major difference between an inverter-driven motor and a vector-controlled motor is that vector control characteristics are not limited to steady-state, but applicable in a dynamic situation like the dc motor.

With vector control, you can obtain dynamic operating characteristics comparable to a dc motor. The major advantage of the vector-driven induction motor is its rugged construction, reliability, low cost and simple construction for high-speed machines. In addition, constant-power operation is somewhat simpler than a separately excited dc motor because it is embedded in the vector control.

Vector control system Implementation

We can present a practical vector control design with Baldor's ASBTS-20 series spindle drive. This system uses indirect vector control to get very accurate speed and positional controllability throughout its wide speed range. Often, this is required in high-performance machine tool spindle applications. Figure 7 shows a functional block diagram of the vector control used in the drive. Except the motor and mounted encoder, all electronics are packaged inside the drive. The drive includes many other functions — handshake logic with NC machine, protective functions and the operator interface etc — but discussions will be limited to vector control.

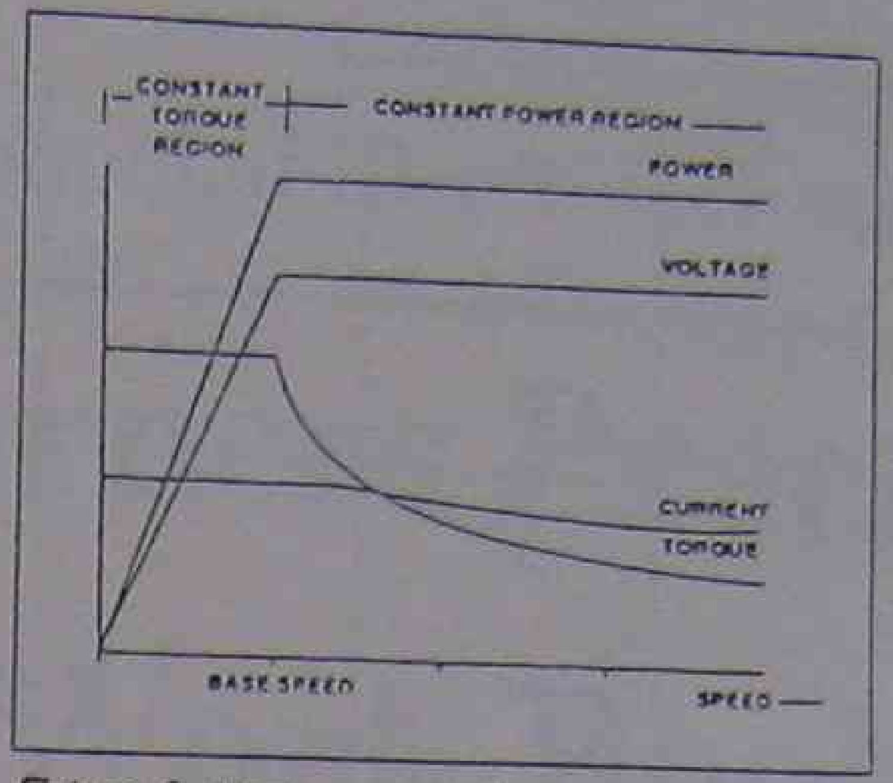

Figure 6. Vector controlled motor characteristics

For vector control, which involves many arithmetic calculations of extended resolution, at least a 16-bit microprocessor, or signal processor is required. The ASBTS-20 uses a NEC 78312 16-bit microcontroller with many functions handled by hardware, either through built-in microcontroller logic or on-board logic.

This effectively relieves software burdens required with fast sampling time.

From the speed command Vc from a host (either CNC or operator), velocity profile, Vr, is generated for every control interrupt cycle. The rate of change in the profile velocity depends on acceleration rate and speed. Vr is then compared to the actual speed of the motor, and the closed-loop compensator determines torque command, Tc, based on the present and past speed errors. Vector control starts from this point. Flux current component Id is calculated from the flux calculator process that simply modifies command flux Idc inversely proportional to the speed above base speed. Because the Tc command is directly proportional to Iq (torque-producing current component), we are ready for vector calculation. From two variables Id and Iq, magnitude and phase of stator current vector can be calculated from equations (11) and (14), and produced slip from equation (12). Rotor flux angle can be estimated by equation (13), so the appropriate arithmetic and digital integration process results in a rotor flux angle of the current interrupt cycle. Based on the magnitude and flux angle, the frame transformation process can be done via a combination of hardware and software. The resulting three phase stator current commands are transformed to an analogue format, filtered and amplified to feed motor current. To eliminate current phase lag due to inductance, a PWM amplifier with high gain current feedback is used.

There are two open-loop parameters that must be tuned properly. These are Idc and Tr (rotor time constant). To optimise performance, these must be tuned correctly with a load test. Because motor parameter Tr varies as motor temperature changes, peak torque capability may be deteriorated as temperature varies. Fortunately, for sizes of most spindle motors, its effect is not so critical in performance as in much larger size motors.

Dr Ohm is with the Baldor motion products group.

For further information contact Australian Baldor, 6 Stanton Rd, Seven Hills NSW 2147.

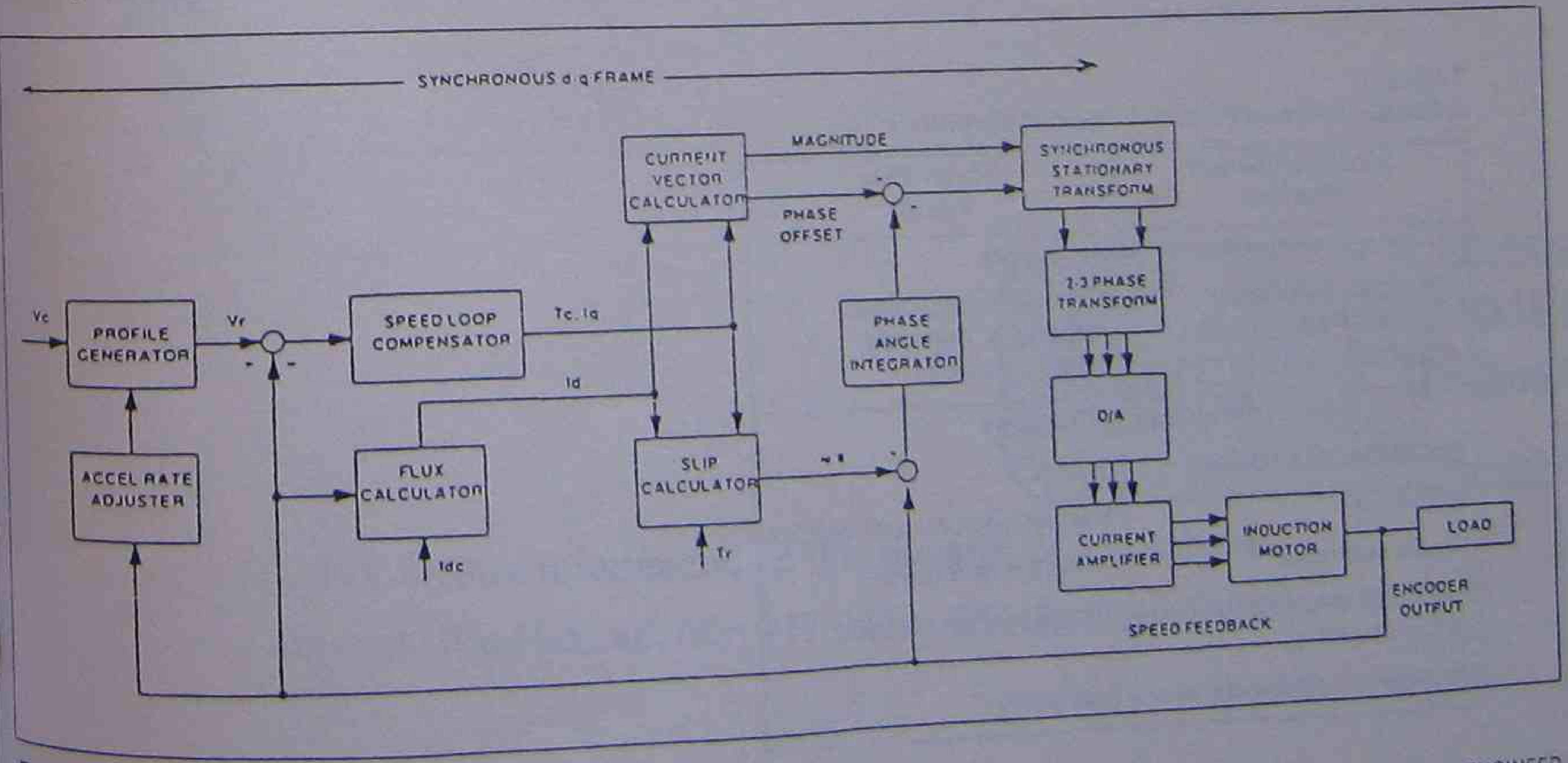

Figure 7. Vector controlled motor drive

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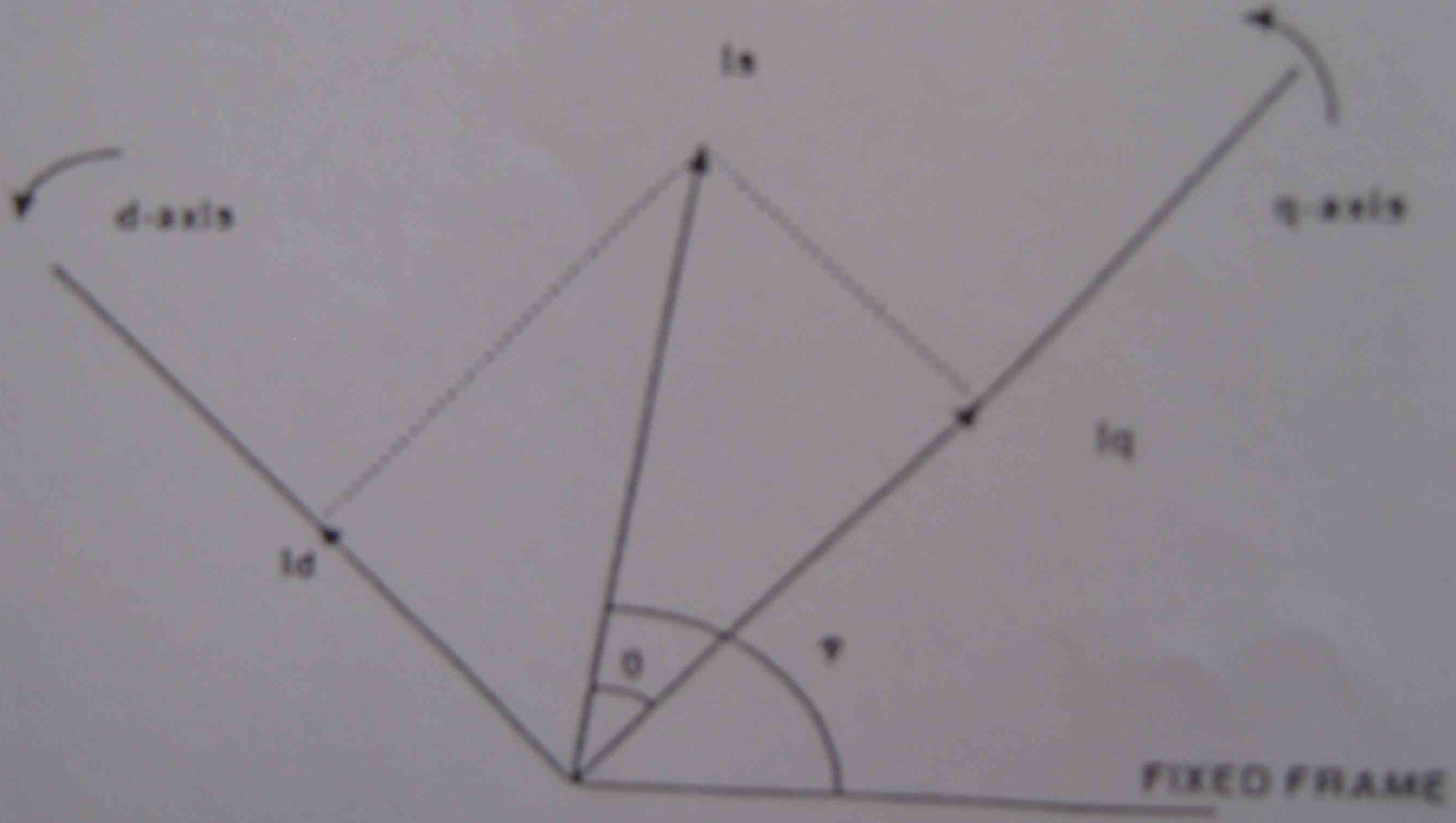

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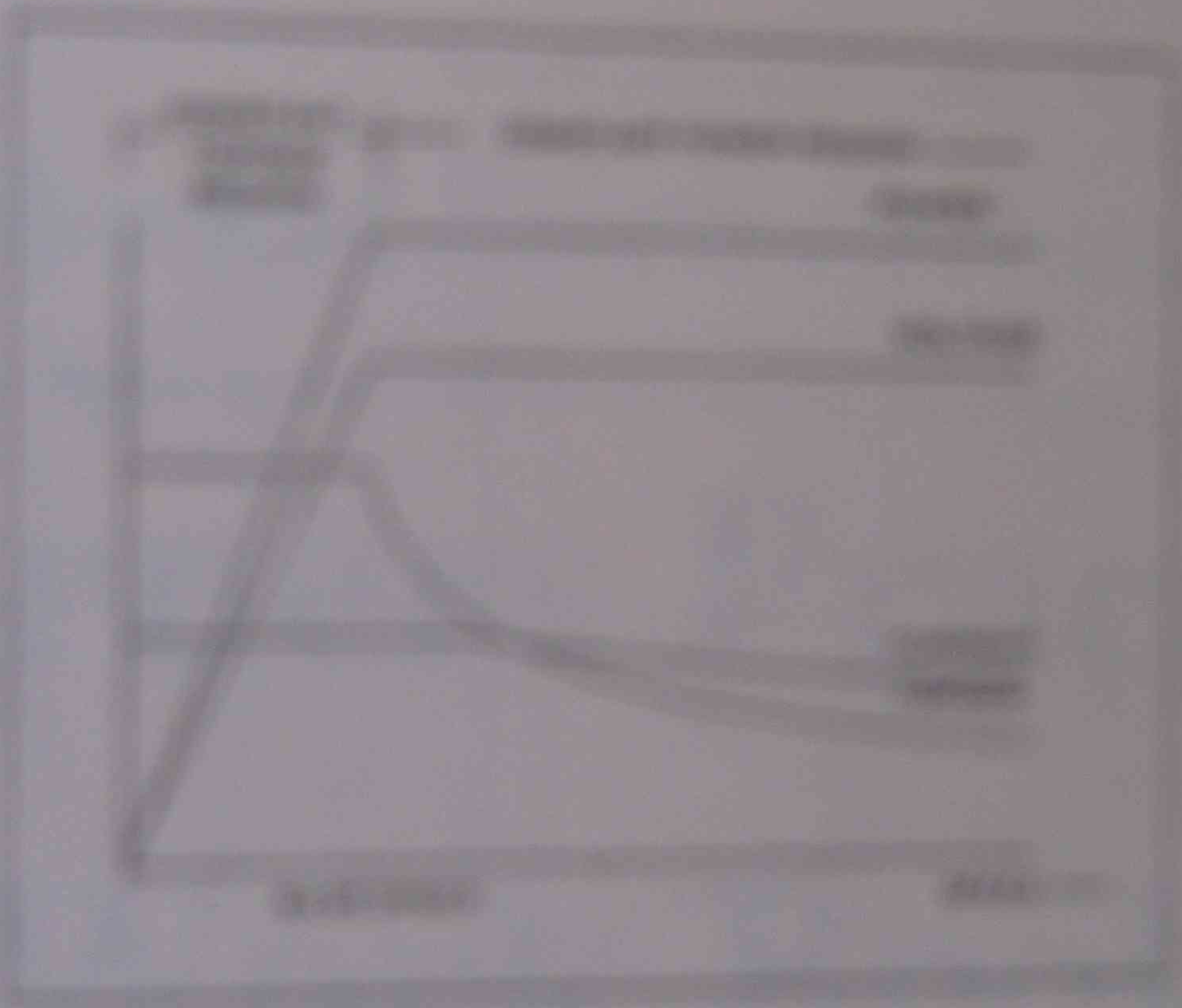

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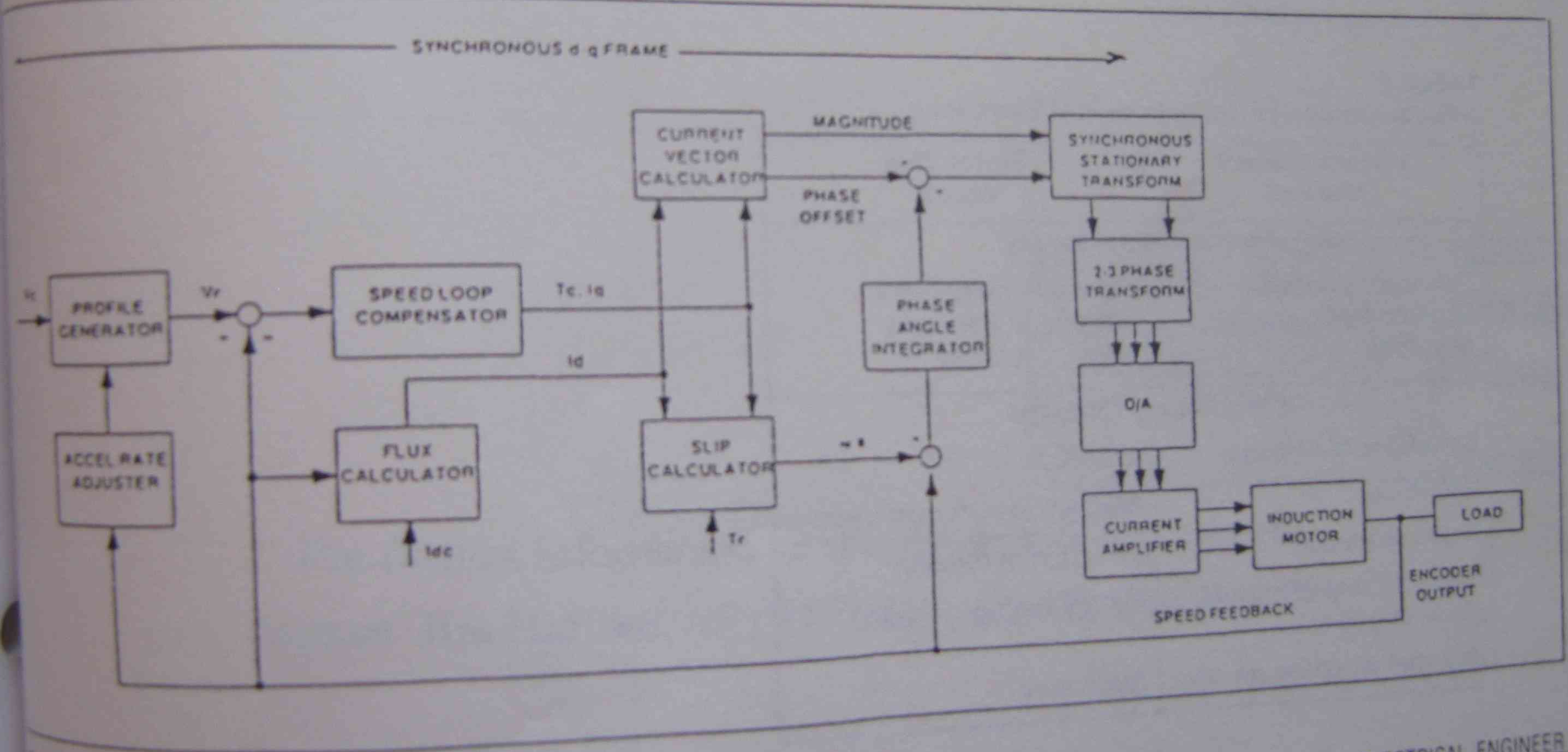

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