ELEVEN

Some Applications

Several applications of power electronics are listed in Art 1.2. Study of all these applications will be a voluminous task. Even then, some of these applications described in this chapter will be of interest to the reader.

11.1. SWITCHED MODE POWER SUPPLY (SMPS)

With advances in electronics, need for dc power supplies for use in integrated circuits (ICs) and digital circuits has increased manifold. For such electronic circuits, NASA was the first to develop a light-weight and compact switched mode power supply in the 1960s for use in its space vehicles. Subsequently, this power supply became popular and presently, annual production of SMPSs may be as high as 70 to 80% of the total number power supplies produced.

At this juncture, a question may arise that controlled dc supply can also be obtained from phase-controlled rectifiers. Then why go in for SMPS ? An ac to dc rectifier operates at supply frequency of 50 (or 60) Hz. In order to obtain almost negligible ripple in the dc output voltage, physical size of filter circuits required is quite large. This makes the dc power supply inefficient, bulky and weighty. On the other hand, SMPS works like a dc chopper. By operating the on/off switch very rapidly, ac ripple frequency rises which can be easily filtered by L and C filter circuits which are small in size and less weighty. It may therefore be inferred that it is the requirement of small physical size and weight that has led to the wide spread use of SMPSs.

As stated above, SMPS is based on the chopper principle. The output dc voltage is controlled by varying the duty cycle of chopper by PWM or FM techniques. The circuit configurations used for SMPS can be classified into four broad categories; namely flyback, pushfull, half bridge and full-bridge.

In SMPS circuits discussed here, PWM technique is used for the inverter. The output of the inverter is then converted to dc by a diode rectifier. As the inverter is made to operate at very high frequency, the ripples on the dc output voltage can be filtered out easily by using small filter components. If the switching devices are power transistors, the chopping frequency is limited to 40 kHz. For power MOSFETs, the chopping frequency is of the order of 200 kHz; as a result, size of the filter circuit and transformer decreases leading to considerable savings. At such high frequency, ferrite core is used in transformers.

The four categories of SMPSs listed above are now discussed briefly.

11.1.1. Flyback Converter

The circuit configuration for flyback converter is shown in Fig. 11.1. It consists of a power MOSFET M1, transformer for isolation purposes, diode D, capacitor C and load. An uncontrolled rectifier converts ac to dc output which is fed to flyback SMPS as shown in Fig. 11.1.

When power MOSFET is turned on, supply voltage V_s is applied to the transformer primary, *i.e.* $v_1 = V_s$. A corresponding voltage v_2, with the polarity as shown in Fig. 11.2(*a*), is induced in the transformer secondary, *i.e.* $v_2 = \frac{V_s}{N_1} N_2$. As v_2 reverse biases diode D, equivalent circuit of Fig. 11.2(*a*) is obtained. Filter capacitance C is assumed large enough so that capacitor voltage $v_c(t)$ = load or output voltage V_0 is taken as almost constant. When M1 is turned off, a voltage of opposite polarity is induced in primary and secondary windings as shown in Fig. 11.2(*b*). Voltage across transformer secondary is $v_2 = -V_0 = -\frac{V_s}{N_1} N_2$. Diode D is forward biased and starts conducting a current i_D. As a result, energy stored in the transformer core is delivered partly to load and partly to charge the capacitor C.

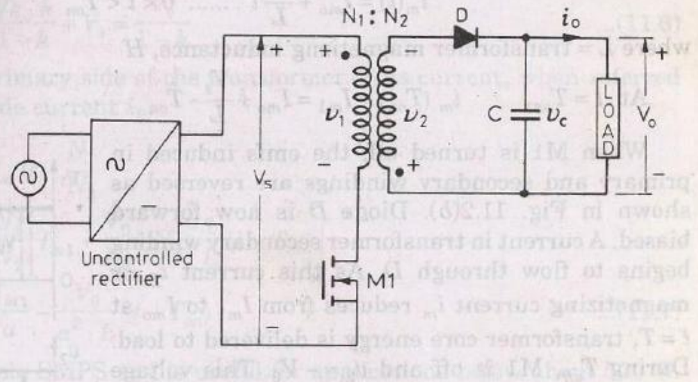


Fig. 11.1. Flyback SMPS.

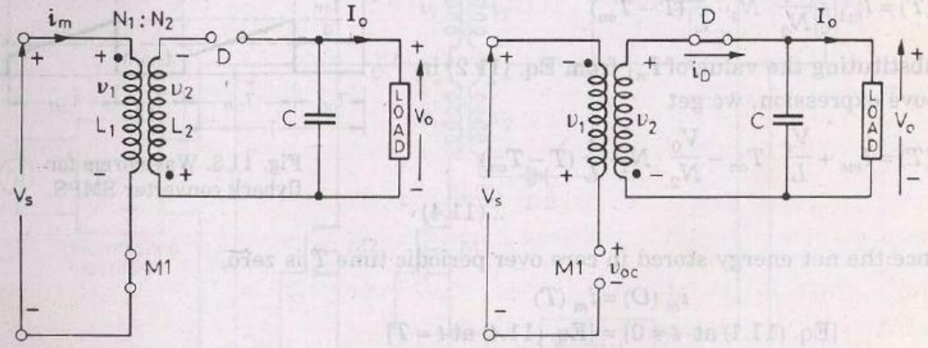


Fig. 11.2. Flyback SMPS equivalent circuit during (*a*) T_{on} and (*b*) T_{off}.

Waveforms for v_1, v_2, transformer magnetizing current i_m and diode current i_D are shown in Fig. 11.3. During the time M1 is on, $v_1 = V_s$, $v_2 = \frac{V_s}{N_1} \cdot N_2$. For magnetizing current, it is assumed that transformer core is not demagnetized completely at the end of periodic time $T = T_{on} + T_{off}$. In other words, it means that transformer magnetizing current at $t = 0$ is not zero but has some positive value I_{mo}. Therefore, during T_{on}, magnetizing current rises linearly from its initial value I_{mo} to I_{m1} at $t = T_{on}$. With the rise of i_m during T_{on}, magnetic energy gets stored in the transformer core. The variation of i_m as shown in Fig. 11.3 can be expressed as under :

$$i_m(t) = I_{mo} + \frac{V_s}{L} t \quad \ldots\ldots \quad 0 < t < T_{on} \qquad \ldots(11.1)$$

where L = transformer magnetizng inductance, H

At $t = T_{on}$, $\quad i_m(T_{on}) = I_{m1} = I_{mo} + \frac{V_s}{L} \cdot T_{on}$ ...(11.2)

When M1 is turned off, the emfs induced in primary and secondary windings are reversed as shown in Fig. 11.2(*b*). Diode *D* is now forward biased. A current in transformer secondary winding begins to flow through *D*. As this current i_D or magnetizing current i_m reduces from I_{m1} to I_{mo} at $t = T$, transformer core energy is delivered to load. During T_{off}, M1 is off and $v_2 = -V_0$. This voltage when referred to primary is $v_1 = -\frac{V_0}{N_2} N_1$. The fall of current i_m during T_{off} can be expressed as under:

$$i_m(t) = I_{m1} - \frac{V_0}{N_2} N_1 \cdot \frac{1}{L}(t - T_{on}) \quad \ldots\ldots \quad T_{on} < t < T$$

...(11.3)

At $t = T$,

$$i_m(T) = I_{m1} - \frac{V_0}{N_2} N_1 \frac{1}{L}(T - T_{on})$$

Substituting the value of I_{m1} from Eq. (11.2) in the above expression, we get

$$i_m(T) = I_{mo} + \frac{V_s}{L} \cdot T_{on} - \frac{V_0}{N_2} \cdot N_1 \cdot \frac{1}{L}(T - T_{on})$$

...(11.4)

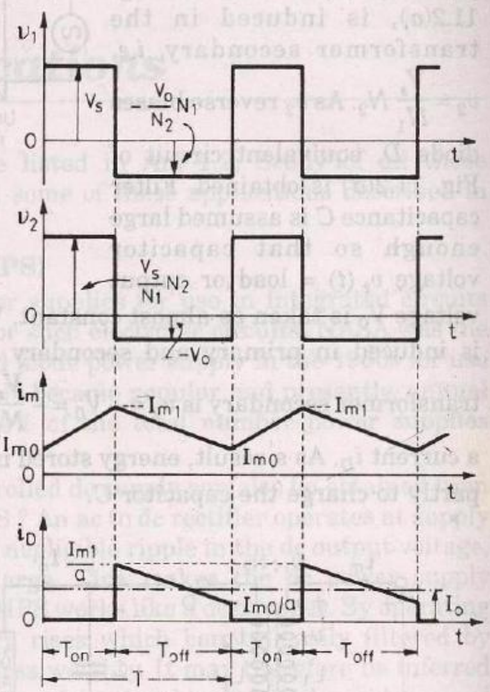


Fig. 11.3. Waveforms for flyback converter SMPS.

Since the net energy stored in core over periodic time T is zero,

$$i_m(O) = i_m(T)$$

or [Eq. (11.1) at $t = 0$] = [Eq. (11.4) at $t = T$]

$$I_{mo} = I_{mo} + \frac{V_s}{L} \cdot T_{on} - \frac{V_0}{N_2} \cdot N_1 \cdot \frac{1}{L}(T - T_{on})$$

or $$V_s \cdot T_{on} = \frac{V_0}{a}(T - T_{on})$$

$\therefore$ Load voltage, $$V_0 = \frac{a \cdot V_s \cdot T_{on}}{T - T_{on}} = \frac{a \cdot V_s \cdot k}{1 - k} \qquad \ldots(11.5)$$

where $a = \frac{N_2}{N_1}$, transformer turns ratio from secondary to primary

and $k = \frac{T_{on}}{T}$, duty cycle of flyback converter.

It is seen from Fig. 11.2 (*b*) that open circuit voltage across M1 is

$$V_{oc} = v_1 + V_s = \frac{V_0}{N_2} \cdot N_1 + V_s = \frac{V_0}{a} + V_s$$

From Eq. (11.5), $$V_{oc} = \frac{V_S \cdot k}{1-k} + V_s = \frac{V_s}{1-k} \quad ...(11.6)$$

Eq. (11.3) gives current on primary side of the transformer. This current, when referred to secondary side, is equal to diode current i_D.

$$\therefore \quad i_D(t) = i_m(t) \cdot \frac{N_1}{N_2}$$

$$= \frac{N_1}{N_2}\left[I_{m1} - \frac{V_0}{N_2} \cdot N_1 \cdot \frac{1}{L}(t - T_{on})\right]$$

$$= \frac{I_{m1}}{a} - \frac{V_0}{a^2 \cdot L}(t - T_{on}) \quad ...(11.7)$$

Flyback converter offers simple SMPS and is useful for applications below about 500 W.

11.1.2. Push-pull Converter

SMPS with push-pull configuration is shown in Fig. 11.4. It uses two power MOSFETs M1 and M2 and a transformer with mid-taps on both primary and secondary sides. As in flyback converter, an uncontrolled rectifier feeds push-pull SMPS. Inductor L and capacitor C are the filter components.

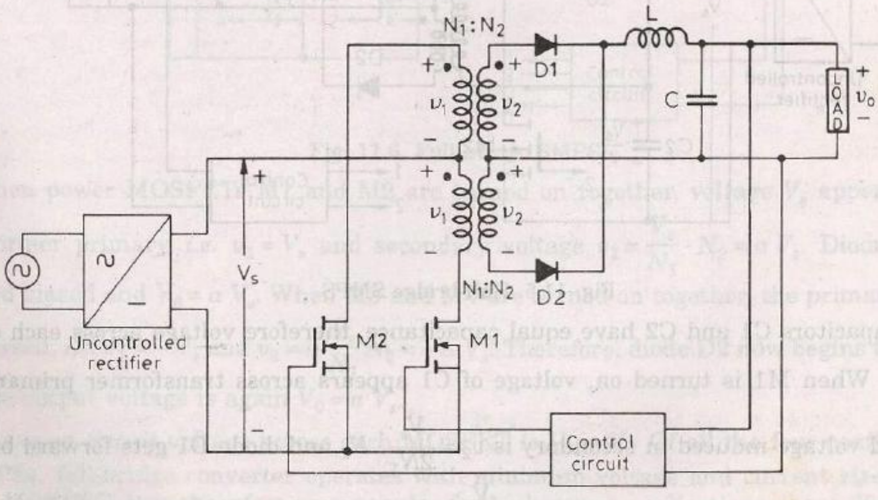


Fig. 11.4. Push-pull SMPS.

When M1 is turned on, V_s is applied to lower half of transformer primary, *i.e.* $v_1 = V_s$. As a result, voltage $v_2 = \frac{V_s}{N_1} N_2$ is induced in both the secondary windings. Voltage v_2 in the upper half secondary forward biases diode D1, therefore load voltage V_0 is given by $V_0 = \frac{V_s}{N_1} N_2 = aV_s$.

When M2 is turned on, $v_1 = -V_s$ is applied to upper half of primary winding. Consequently, $v_2 = -\frac{V_s}{N_1} N_2$ is induced in both the transformer secondaries. As v_2 is negative, diode D2 gets forward biased and $V_0 = a\ V_s$ as before.

This shows that voltage on primary swings from $+V_s$ with M1 on to $-V_s$ with M2 on. Power MOSFETs M1 and M2 operate with duty cycle of 0.5. When M1 is off, the voltage across M1 terminals is $V_{oc} = 2V_s$. As both M1 and M2 are subjected to open-circuit voltage of $2V_s$, this configuration is suitable for low-voltage applications only.

11.1.3. Half-bridge Converter

The circuit for half-bridge SMPS configuration is shown in Fig. 11.5. It consists of an uncontrolled rectifier, two capacitors C1 and C2, two power MOSFETs M1 and M2, one transformer with mid-tap on the secondary side, two diodes D1 and D2 and filter components L and C.

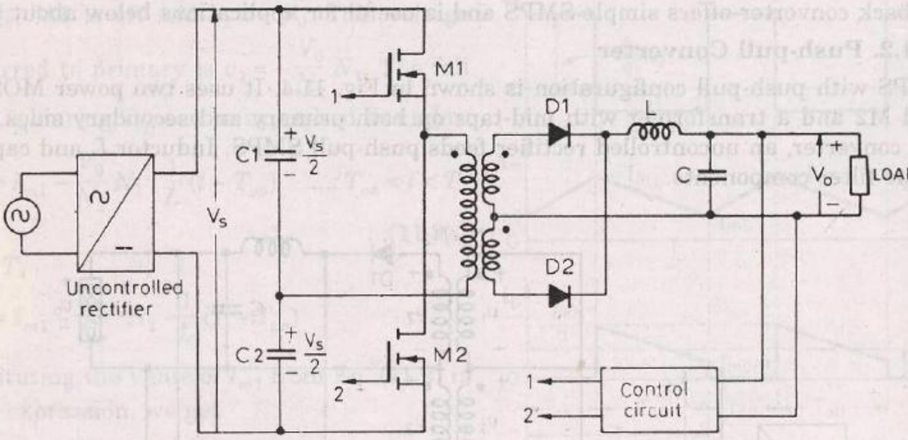


Fig. 11.5. Half-bridge SMPS.

Two capacitors C1 and C2 have equal capacitance, therefore voltage across each of the two is $\frac{V_s}{2}$. When M1 is turned on, voltage of C1 appears across transformer primary, *i.e.* $v_1 = \frac{V_s}{2}$ and voltage induced in secondary is $v_2 = \frac{V_s}{2N_1} \cdot N_2$ and diode D1 gets forward biased.

When M2 is turned on, a reverse voltage of $\frac{V_s}{2}$ appears across transformer primary from C2, *i.e.* $v_1 = -\frac{V_s}{2}$ and voltage induced in secondary winding is $v_2 = -\frac{V_s}{2N_1} N_2$, therefore diode D2 gets forward biased. This means that transformer primary voltage swings from $-\frac{V_s}{2}$ to $+\frac{V_s}{2}$. Average output voltage, however, is

$$V_0 = \frac{V_s}{2N_1} \cdot N_2 = 0.5\, a\, V_s$$

When M1 is off, open circuit voltage across M1 terminals is $V_{oc} = V_s$. When M2 is off, as before $V_{oc} = V_s$. For *h.v. dc* applications, half-bridge converter is, therefore, preferred over push-pull converters. For *l.v. dc* applications, push-pull SMPS is preferred due to low MOSFET currents.

11.1.4. Full-bridge Converter

The circuit arrangement for a full-bridge SMPS is shown in Fig. 11.6. It consists of an uncontrolled rectifier, four power MOSFETs, transformer with mid- tap secondary, two diodes and *LC* filter circuit. As in all the previous circuits, the function of control circuit is to sense the output load voltage and to decide about the duty ratio of MOSFETs.

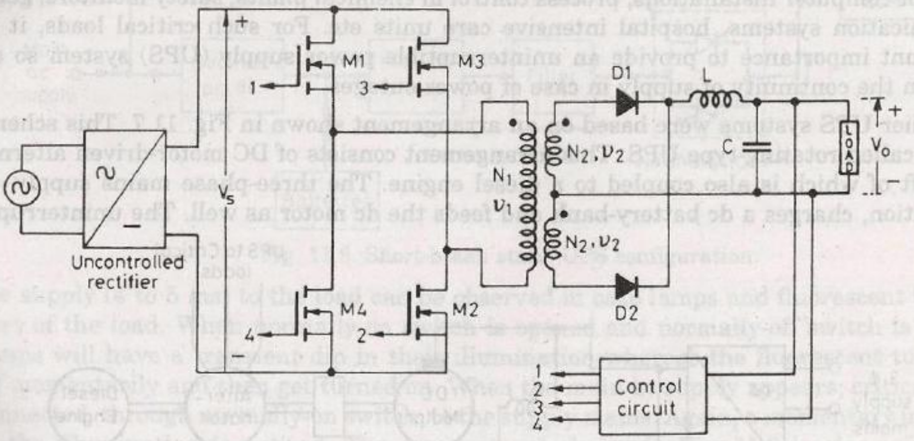


Fig. 11.6. Full bridge SMPS.

When power MOSFETs M1 and M2 are turned on together, voltage V_s appears across transformer primary, *i.e.* $v_1 = V_s$ and secondary voltage $v_2 = \frac{V_s}{N_1} \cdot N_2 = a\, V_s$. Diode D1 gets forward biased and $V_0 = a\, V_s$. When M3 and M4 are turned on together, the primary voltage is reversed, *i.e.* $v_1 = -V_s$ and $v_2 = -\frac{V_s}{N_1} N_2 = -a\, V_s$. Therefore, diode D2 now begins to conduct and the output voltage is again $V_0 = a\, V_s$.

The open circuit voltage across each MOSFET is $V_{oc} = V_s$. Of all the four configurations of SMPSs, full-bridge converter operates with minimum voltage and current stress on the power MOSFET. It is therefore very popular for high power applications above 750 W.

The overall size of SMPSs is dependent on its operating frequency. Use of power transistors is limited to approximately 40 to 50 kHz. Above this operating frequency, power MOSFETs are used up to about 200 kHz.

The main advantages of SMPSs over conventional linear power supplies are as under :

(*i*) For the same power rating, SMPS is of smaller size, lighter in weight and possesses higher efficiency because of its high-frequency operation.

(*ii*) SMPS is less sensitive to input voltage variations.

The disadvantages of SMPS are as under :

(*i*) SMPS has higher output ripple and its regulation is worse

(*ii*) SMPS is a source of both electromagnetic and radio interference due to high frequency switching.

(*iii*) Control of radio frequency noise requires the use of filters on both input and output of SMPS.

The advantages possessed by SMPSs far outweigh their shortcomings. This is the reason for their wide-spread popularity and growth.

11.2. UNINTERRUPTIBLE POWER SUPPLIES

There are several applications where even a temporary power failure can cause a great deal of public inconvenience leading to large economic losses. Examples of such applications are major computer installations, process control in chemical plants, safety monitors, general communication systems, hospital intensive care units etc. For such critical loads, it is of paramount importance to provide an uninterruptible power supply (UPS) system so as to maintain the continuity of supply in case of power outages.

Earlier UPS systems were based on an arrangement shown in Fig. 11.7. This scheme is usually called rotating-type UPS. This arrangement consists of DC motor-driven alternator, the shaft of which is also coupled to a diesel engine. The three-phase mains supply, after rectification, charges a dc battery-bank and feeds the dc motor as well. The uninterruptible

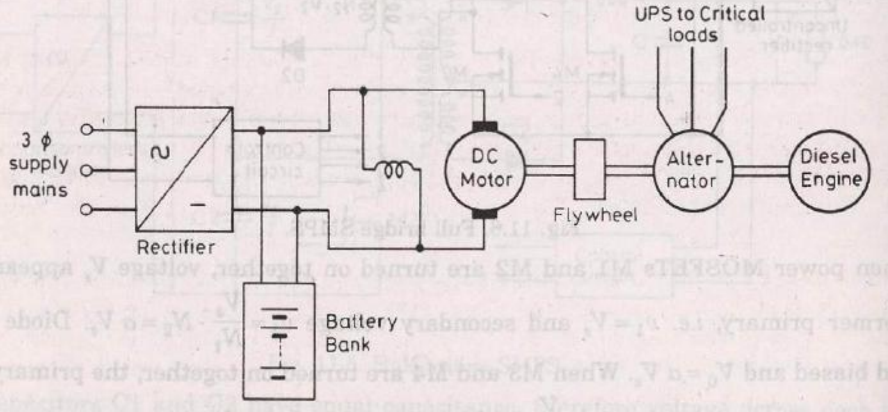


Fig. 11.7. Rotating-type UPS system based on dc motor/alternator set.

power supply needed is taken from the alternator output terminals. When mains supply fails, the diesel engine is run to take over the load. Starting of the diesel engine takes 10 to 15 seconds. During this period, the battery-bank is able to maintain the alternator speed through the dc motor and the flywheel, thus giving a no-break supply to the critical load. At present, however, static UPS systems are becoming popular up to a few kVA ratings.

Static UPS systems are of two types ; namely short-break UPS and no-break UPS. In short-break UPS, the load gets disconnected from the power source for a short duration of the order of 4 to 5 ms. In no-break UPS, load gets continuous uninterrupted supply from the power source. These are now discussed briefly.

Short-break UPS. In situations where short interruption (4 to 5 ms) in supply can be tolerated, the short-break UPS shown in Fig. 11.8 is used. In this system, main ac supply is rectified to dc. This dc output from the rectifier charges the batteries and is also converted to ac by an inverter, Fig. 11.8. After passing through the filter, ac can be delivered to load in case normally-off contacts are closed. Under normal circumstances, normally-on contacts are closed and normally-off contacts are open and the main supply delivers ac power to the load. At the same time, the rectifier supplies continuous trickle charge to batteries to keep them fully charged. In the event of power outage, normally-off switch is turned-on and the batteries deliver ac power to critical load through the inverter and filter. A momentary interruption in

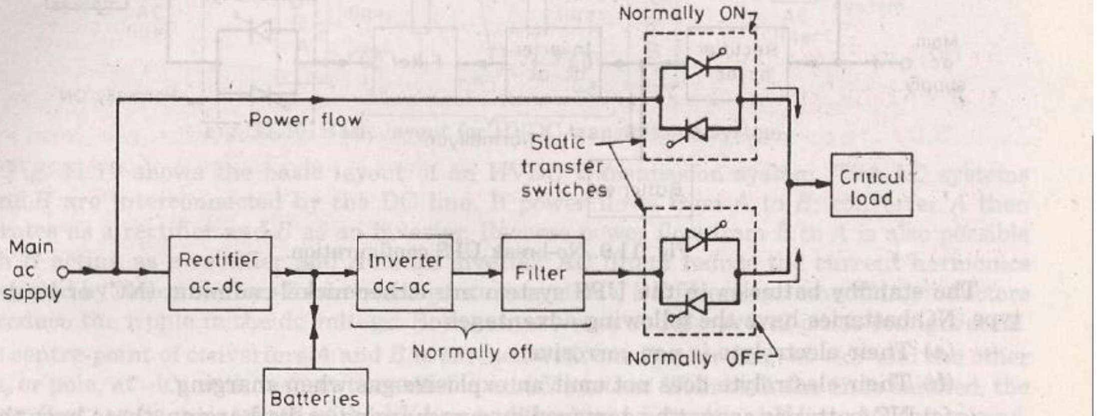


Fig. 11.8. Short-break static UPS configuration.

the supply (4 to 5 ms) to the load can be observed in case lamps and fluorescent tubes are a part of the load. When normally-on switch is opened and normally-off switch is turned on, lamps will have a transient dip in their illumination whereas the fluorescent tubes will be off momentarily and then get turned on. When the main ac supply appears, critical load gets connected, through normally-on switch, to the supply mains. Again, a momentary interruption in the illumination is noticed. The arrangement shown in Fig. 11.8 is also referred to as *stand-by power supply*.

No-break UPS. When a no-break supply is required, the static UPS system shown in Fig. 11.9 is used. In this system, main ac supply is rectified and the rectifier delivers power to maintain required charge on the batteries. Rectifier also supplies power to inverter continuously which is then given to ac-type load through filter and normally-on switch. In case of main-supply failure, batteries at once take over with no-break of supply to the critical load. No dip or discontinuity in the illumination is observed in case of no-break UPS. This configuration of Fig. 11.9 has the following additional advantages :

(*i*) The inverter can be used to condition the supply delivered to load.
(*ii*) Load gets protected from transients in the main ac supply.
(*iii*) Inverter output frequency can be maintained at the desired value.

In case inverter failure is detected, the load is switched on to the main ac supply directly by turning on the normally-off static switch and opening the normally-on static switch. The transfer of load from inverter to main ac supply takes 4 to 5 ms by static transfer switch as compared to 40 to 50 ms for a mechanical contactor. After inverter fault is cleared, uninterruptible power supply is again restored to the load through the normally on switch. The batteries are now recharged from the main supply by adjusting the charger at maximum charge rate so that batteries are charged to their full capacity in the shortest possible time.

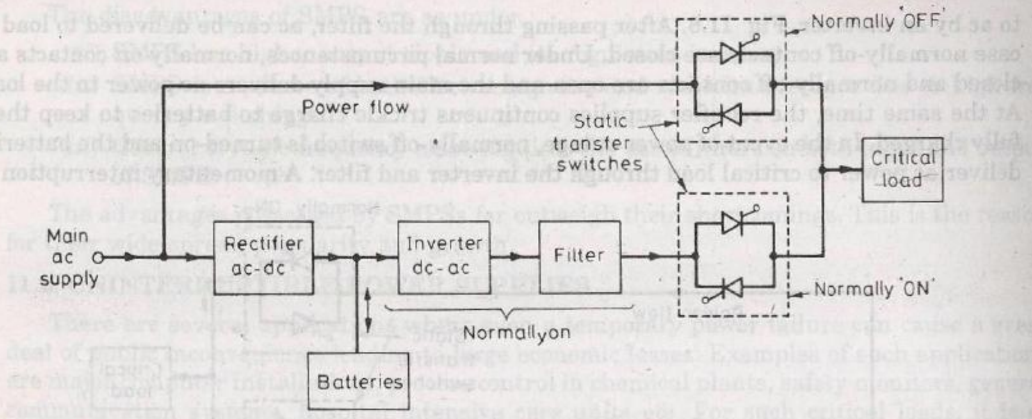


Fig. 11.9. No-break UPS configuration.

The standby batteries in the UPS system are either nickel-cadmium (NC) or lead-acid type. NC batteries have the following advantages :

(*a*) Their electrolyte is non-corrosive.

(*b*) Their electrolyte does not emit an explosive gas when charging.

(*c*) NC batteries cannot be damaged by overcharging or discharging, these have therefore longer life.

Cost of NC batteries is, however, two or three times that of lead-acid batteries.

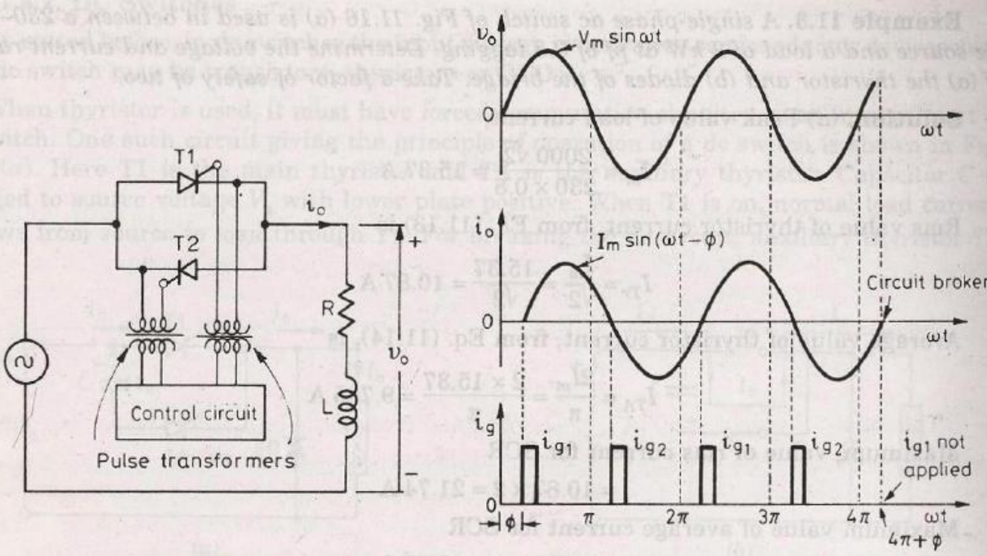
The time period for which a battery or a battery-bank can deliver power to load through inverter at the required voltage level depends upon (*i*) the size of the batteries and (*ii*) nature of the load.

11.5. STATIC CIRCUIT BREAKERS

Static circuit breakers are semiconductor-based circuits capable of providing a fast and reliable interruption to a continuous current. Static circuit breakers are of two types; static ac circuit breakers and static dc circuit breakers. High-current circuit breakers employing thyristors are now discussed briefly.

11.5.1. Static AC Circuit Breakers

Static ac switch can be made to operate as a static ac circuit breaker. In Fig. 11.19 (*a*) is shown a simplified circuit configuration for static ac circuit breaker and Fig. 11.19 (*b*) gives relevant voltage and current waveforms. As in static switches, thyristors 1 and 2 in Fig. 11.19(*a*) are turned on at the instant load current is passing through zero. For breaking the circuit, the triggering pulse is withdrawn. For example, at $\omega t = 4\pi + \phi$, if triggering pulse i_{g1} is not applied to T1, it will not get turned on. T2 is already off just before $\omega t = 4\pi + \phi$. Therefore, the continuity of the circuit is broken. So when turn-off command is received by the control circuit due to some system fault, the gating pulse is withdrawn from T1 or T2 and eventually



(*a*) (*b*)

Fig. 11.19. (*a*) Static *ac* circuit breaker and (*b*) its relevant voltage and current waveforms.

the circuit is broken. In case turn-off command is received just after $3\pi + \phi$, load current will be broken only at $4\pi + \phi$, *i.e.* a delay of π radians or half-cycle is a must. If turn-off command is received at the instant $(3\pi + \phi) < \omega t < (4\pi + \phi)$, even then the circuit is broken at the instant $\omega t = 4\pi + \phi$ only. This shows that maximum time delay for breaking the circuit is one half-cycle *i.e.* $\frac{\pi}{\omega}$ seconds after turn-off command is accepted by the control circuit due to some exigencies in the system.

11.5.2. Static DC Circuit Breakers

A simple arrangement of static dc circuit breaker is shown in Fig. 11.20. This circuit is similar to that shown in Fig. 5.4(*a*), pertaining to class-C commutation. As stated before, when input voltage to a circuit consisting of thyristors is dc, forced commutation is essential for turning off a thyristor. For complete analysis, refer to section 5.3. Here only brief discussion is given.

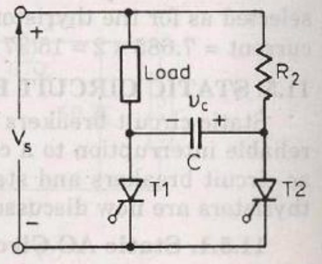


Fig. 11.20. Static dc circuit breaker.

When main thyristor T1 is turned on, load voltage becomes equal to source voltage V_s and capacitor C begins to charge through the circuit V_s, R_2, C and T1. Eventually capacitor C gets charged with right hand plate positive. For breaking the circuit, auxiliary thyristor T2 is turned on. Capacitor voltage v_c at once applies a reverse voltage V_s across SCR T1 and turns it off. After T1 is force commutated, capacitor will charge from $+V_s$ to $-V_s$ through the circuit V_s, load, C and T2. When C is fully charged to $-V_s$ (left hand plate positive), current through load will be zero and at the same time current through R_2 will be less than the holding current of SCR T2. As a result, T2 will get turned off naturally. From this, the value of R_2 can be determined.

11.6. SOLID STATE RELAYS

AC and dc static switches can be used as solid state relays (SSRs) in ac and dc circuits respectively. In ac circuits, thyristors or triacs are used whereas in dc circuits transistors are preferred. Solid state relays have no contacts or moving parts. These are now being used extensively and are replacing the conventional contact-type electromagnetic relays in applications like control of motor drives, resistance heating etc. SSRs need electrical isolation between control circuit and the load circuit by means of optocouplers or pulse transformers.

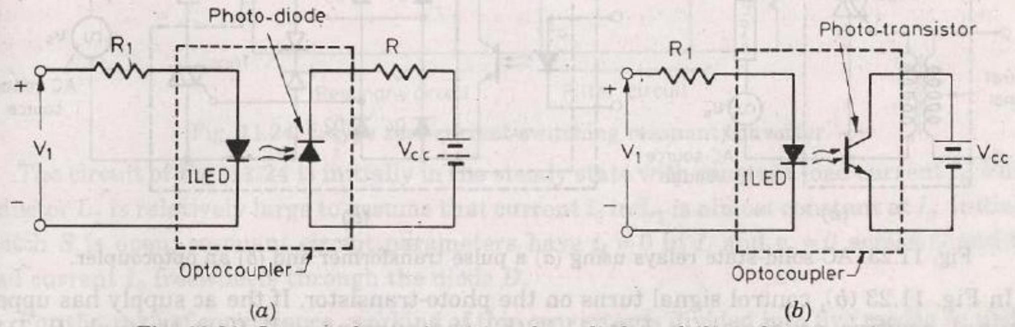


Fig. 11.21. Optocouplers using (*a*) a photo-diode and (*b*) a photo-transister.

An optocoupler consists of infra-red light emitting diode (ILED) and a photo-diode or a photo-transistor. An optocoupler having ILED and photo-diode is shown in Fig. 11.21 (*a*). A short pulse V_1 applied to ILED will cause it to emit light on to photo-diode which will then begin to conduct in the reverse direction as shown. An optocoupler using photo-transistor is shown in Fig. 11.21 (*b*). As before, a short pulse V_1 applied to ILED will throw light on the base of photo-transistor and turn it on. As photo-transistor is more sensitive than a photo-diode, optocouplers based on opto-transistors are more common.

11.6.1. DC Solid State Relays

A dc solid state relay using opto-coupler for isolation purposes is shown in Fig. 11.22. When control pulse V_C is applied to ILED, it emits light and turns on the photo-transistor. The current output from the photo-transistor acts as the base current for transistor *TR*. Consequently *TR* is turned on and source voltage V_s is applied to load.

When control pulse V_C is absent, *TR* gets turned off and load voltage is zero.

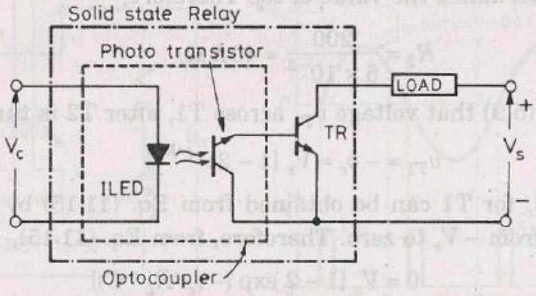


Fig. 11.22. DC solid-state relay using an optocoupler.

11.6.2. AC Solid State Relays

Fig. 11.23 shows two basic circuits for ac solid-state relays. Fig. 11.23 (*a*) uses a pulse transformer for isolation purposes and in Fig. 11.23 (*b*), isolation is provided by an optocoupler. When control signal appears across the primary of pulse transformer, its secondary applies a triggering pulse to turn on the triac. As a result, circuit is completed through v_s, load and triac and therefore, source voltage is applied to the load.

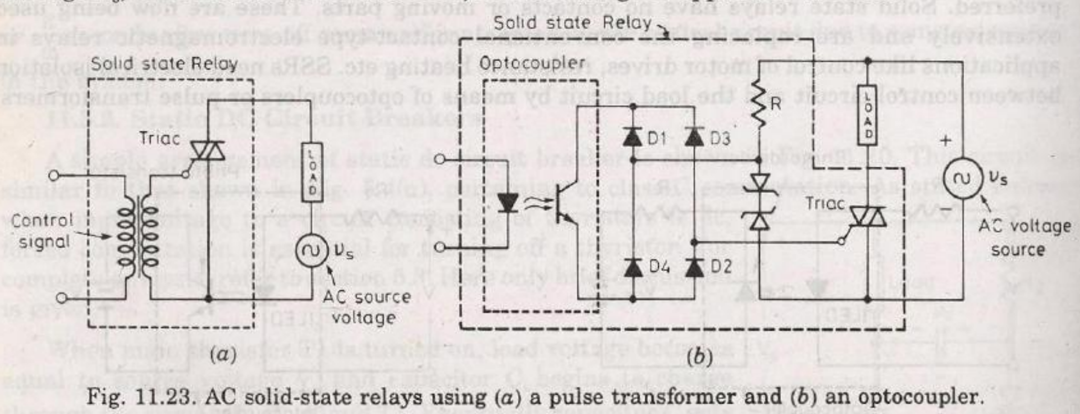


Fig. 11.23. AC solid-state relays using (*a*) a pulse transformer and (*b*) an optocoupler.

In Fig. 11.23 (*b*), control signal turns on the photo-transistor. If the ac supply has upper terminal positive as shown in Fig. 11.23(*b*), the current will flow through *R*, D1, photo-transistor, D2, triac gate and source. This current will turn on the triac and load gets energised by source voltage v_s. The function of *R* is to limit the flow of gate current of triac. If lower terminal of ac supply is positive, the current will flow through triac gate, D3, photo-transistor, D4, *R* and source v_s. Triac gets turned on and source voltage is applied to load.

11.7. RESONANT CONVERTERS

In SMPSs discussed in Art. 11.1 and in the PWM inverters described in Chapter 8, the switching devices are made to turn-on and turn-off the entire load current at high *di/dt*. The devices handling high di/dt also experience high-voltage stresses across them; due to these two effects, there are increased power losses in the switching devices. In case size and weight of the converter components is to be reduced, switching frequencies are increased. At these high frequencies, switching losses and high-voltage stresses are further aggravated. Another major drawback of high di/dt and high dv/dt caused by rapid on and off of the switching devices is the electromagnetic interference.

The shortcomings enunciated above can be minimised if each switch in a converter is turned on and off when the voltage across it and/or current through it is zero at the instant of switching. The converter circuits which employ zero-voltage and/or zero-current switching are called *resonant converters*. In most of these converters, some form of *L-C* resonance is used, that is why these are known as resonant converters.

In this section, resonant converters employing zero-current switching (ZCS) and zero-voltage switching (ZVS) are described.

11.7.1. Zero-Current Switching Resonant Converters

There are two types of ZCS resonant converters, *L*-type and *M*-type. Both of these circuit topologies use L and C as a series resonant circuit; in addition L also limits di/dt of the switching current. Here first *L*-type and then *M*-type ZCS resonant converters are presented.

11.7.1.1. L-type ZCS Resonant Converters. An *L*-type ZCS resonant converter is shown in Fig. 11.24. The switching device S in the figure can be a GTO, thyristor, BJT, power MOSFET or IGBT. At low kilohertz range; GTO, thyristor, transistor or IGBT is used whereas for megahertz range, power MOSFETs are preferred. Inductor L and capacitor C near the dc source V_s form a resonant circuit whereas L_1, C_1 near the load constitute a filter circuit. Direction of currents and plarities of voltages as marked in Fig. 11.24 are treated as positive.

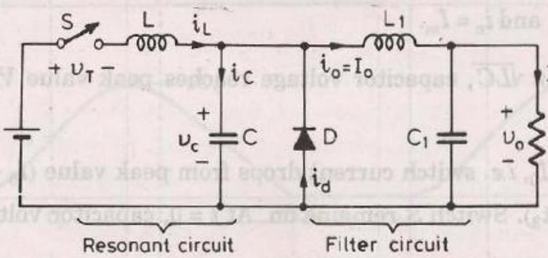


Fig. 11.24. *L*-type zero-current-switching resonant converter.

The circuit of Fig. 11.24 is initially in the steady state with constant load current I_0. Filter inductor L_1 is relatively large to assume that current i_0 in L_1 is almost constant at I_0. Initially, switch S is open; resonant circuit parameters have $i_L = 0$ in L and $v_c = 0$ across C and the load current I_0 freewheels through the diode D.

For the sake of convinience, working of this converter is divided into five modes as under. For all these modes, time t is taken as zero at the beginning of each mode.

Mode I. ($0 \le t \le t_1$). At $t = 0$, switch S is turned on. As I_0 is freewheeling through diode D, voltage across ideal diode $v_D = 0$ and also $v_C = 0$, Fig. 11.25(a). It implies that source voltage V_s gets applied across L and the switch current i_L begins to flow through V_s, switch S, L and diode D, Fig. 11.25(a). Therefore, $V_s = L\, di/dt$. It gives $i_L = \frac{V_s}{L} t$. It shows that inductor or switch current i_L rises linearly from its zero initial value. The diode current i_D is given by

$$i_D = I_0 - i_L = I_0 - \frac{V_s}{L} t \qquad ...(11.15)$$

At $t = t_1$, $\quad i_L = \frac{V_s}{L} \cdot t_1 = I_0$. This gives

$$t_1 = \frac{I_0 \cdot L}{V_s}$$

Also, at $t = t_1$, $i_D = I_0 - I_0 = 0$. Soon after t_1, as i_D tends to reverse, diode D gets turned off. As a result of this, short circuit across C is removed.

Mode II ($0 \le t \le t_2$). Switch S remains on. As D turns off at $t = 0$, current I_0 flows through V_s, L, L_1 and R. In Figs. 11.25(a) and (b), constant current through L_1 and R is represented by current source I_0. Also, a current i_C begins to build up through resonant circuit consisting of V_s, L and C in series. The inductor current i_L is, therefore, given by

$$i_L = I_0 + i_C = I_0 + I_m \sin \omega_0 t \qquad ...(11.16)$$

where $I_m = V_s \sqrt{\frac{C}{L}} = \frac{V_s}{Z_0}$ and $\omega_0 = \frac{1}{\sqrt{LC}}$. Here $Z_0 = \sqrt{\frac{L}{C}}$ is the characteristic impedance of the resonant circuit.

The capacitor current is $i_c = I_m \sin \omega_0 t$ and capacitor voltage v_c is given by

$$v_c(t) = V_s (1 - \cos \omega_0 t) \qquad ...(11.17)$$

The peak value of current i_L is $I_p = I_0 + I_m$ and it occurs at $t = \frac{\pi}{2\omega_0} = \frac{\pi}{2}\sqrt{LC}$. At this instant, $v_c = V_s\,[1 - \cos \pi] = 2V_s$ and $i_c = I_m$.

When $t = t_2 = \frac{\pi}{\omega_0} = \pi\sqrt{LC}$, capacitor voltage reaches peak value $V_{cp} = V_s\,[1 - \cos \pi] = 2V_s$ and $i_c = 0$.

Also, at $t = t_2$, $i_L = I_0$, *i.e.* switch current drops from peak value $(I_0 + I_m)$ to I_0.

Mode III ($0 \le t \le t_3$). Switch S remains on. At $t = 0$, capacitor voltage is $2V_s$. As i_c tends to reverse at $t = 0$, capacitor begins to discharge and force a current $i_c = V_s \sqrt{\frac{C}{L}} \sin \omega_0 t$ opposite to i_L, Fig. 11.25(c), so that inductor or device current i_L is given by

$$i_L = I_0 - i_c = I_0 - I_m \sin \omega_0 t \qquad ...(11.18)$$

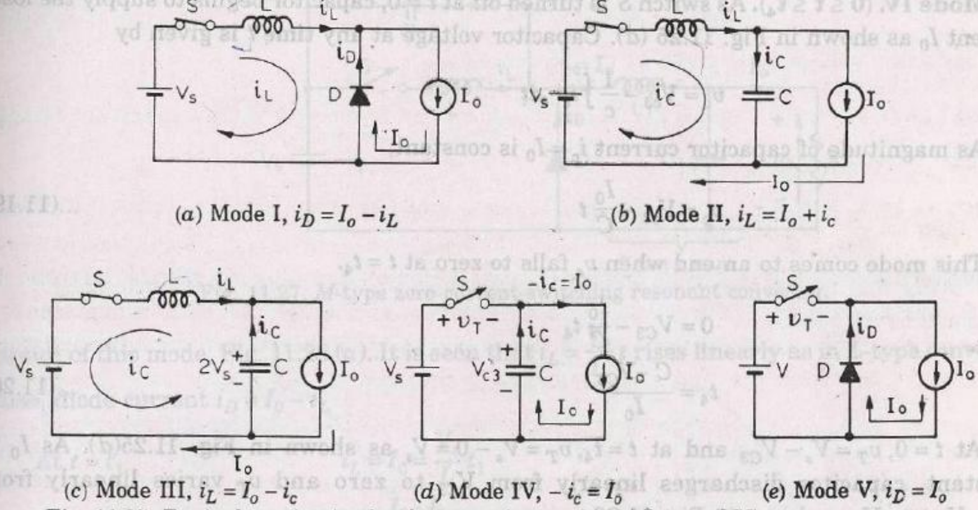
and capacitor voltage $v_c = 2V_s \cos \omega_0 t$. Current i_L falls to zero when $t = t_3$,

i.e. $\quad i_L = 0 = I_0 - I_m \sin \omega_0 t_3$

or $\quad t_3 = \sqrt{LC} \sin^{-1}(I_0/I_m)$

At $t = t_3$, $\quad v_c = 2V_s \cos \omega_0 t_3 = 2V_s \left[\frac{\sqrt{I_m^2 - I_0^2}}{I_m}\right] = V_{c3}$

During this mode, $i_c = I_m \sin \omega_0 t$ and as i_L falls to zero at t_3, switching device S gets turned off. Note that current i_c in this mode flows opposite to its positive direction, it is therefore shown negative in Fig. 11.25(c) and in Fig. 11.26. At $t = t_3$, the value of $i_c = -I_0$.



(a) Mode I, $i_D = I_o - i_L$ (b) Mode II, $i_L = I_o + i_c$

(c) Mode III, $i_L = I_o - i_c$ (d) Mode IV, $-i_c = I_0$ (e) Mode V, $i_D = I_o$

Fig. 11.25. Equivalent circuits for the operating modes of *L*-type ZCS resonant converter.

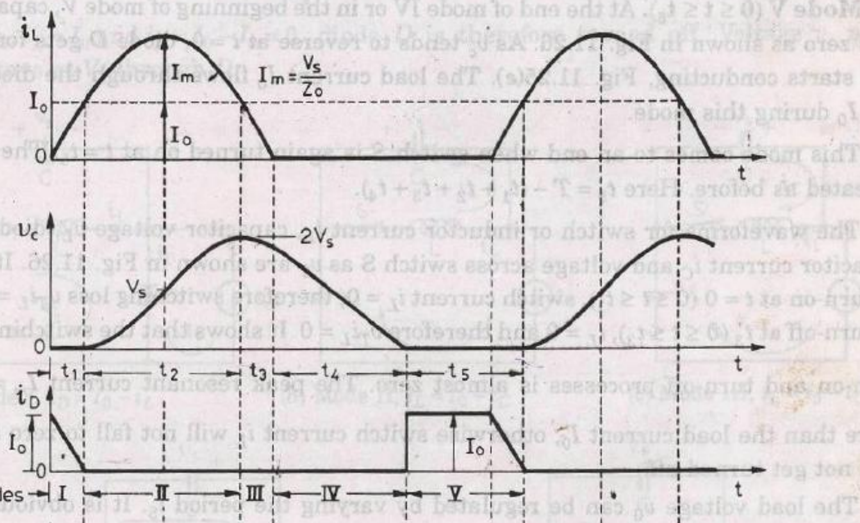


Fig. 11.26. Waveforms for *L*-type ZCS resonant converter.

Mode IV. ($0 \le t \le t_4$). As switch S is turned off at $t = 0$, capacitor begins to supply the load current I_0 as shown in Fig. 11.25 (*d*). Capacitor voltage at any time t is given by

$$v_c = V_{c3} - \frac{1}{c}\int i_c \cdot dt$$

As magnitude of capacitor current $i_c = I_0$ is constant,

$$v_c = V_{c3} - \frac{I_0}{C}t \qquad \text{...(11.19)}$$

This mode comes to an end when v_c falls to zero at $t = t_4$.

or $$0 = V_{C3} - \frac{I_0}{C}t_4$$

or $$t_4 = \frac{C \cdot V_{C3}}{I_0} \qquad \text{...(11.20)}$$

At $t = 0$, $v_T = V_s - V_{C3}$ and at $t = t_4$, $v_T = V_s - 0 = V_s$ as shown in Fig. 11.25(*d*). As I_0 is constant, capacitor discharges linearly from V_{C3} to zero and v_T varies linearly from $(V_s - V_{C3})$ to V_s as shown in Fig. 11.26.

Mode V ($0 \le t \le t_5$). At the end of mode IV or in the beginning of mode V, capacitor voltage v_c is zero as shown in Fig. 11.26. As v_c tends to reverse at $t = 0$, diode D gets forward biased and starts conducting, Fig. 11.25(*e*). The load current I_0 flows through the diode D so that $i_D = I_0$ during this mode.

This mode comes to an end when switch S is again turned on at $t = t_5$. The cycle is now repeated as before. Here $t_5 = T - (t_1 + t_2 + t_3 + t_4)$.

The waveforms for switch or inductor current i_L, capacitor voltage v_L, diode current i_D, capacitor current i_C and voltage across switch S as v_T are shown in Fig. 11.26. It is seen that at turn-on at $t = 0$ ($0 \le t \le t_1$), switch current $i_L = 0$, therefore switching loss $v_T i_L = 0$. Similarly, at turn-off at t_3 ($0 \le t \le t_3$), $i_L = 0$ and therefore $v_T i_L = 0$. It shows that the switching loss during turn-on and turn-off processes is almost zero. The peak resonant current $I_m = \frac{V_s}{Z_0}$ must be more than the load current I_0, otherwise switch current i_L will not fall to zero and switch S will not get turned off.

The load voltage v_0 can be regulated by varying the period t_5. It is obvious that longer the period t_5, lower is the load voltage.

11.7.1.2. M-type ZCS Resonant Converter. An *M*-type ZCS resonant converter is shown in Fig. 11.27. As before, L and C form the resonant circuit and L_1, C_1 the filter circuit. Capacitor C is connected across the series combination of switch S and L; but in L-type converter, C is connected across diode D. Working of this converter can be divided into five modes as for the L-type resonant converter. The time origin $t = 0$ is redefined at the beginning of each mode.

Mode I ($0 \le t \le t_1$). Prior to mode *I*, *i.e.* before switch S is turned on at $t = 0$, load current I_0 freewheels through diode D. Also, voltage $v_c = V_s$ before S is closed. At $t = 0$, switch S is turned on. Now current i_L begins to develop through V_s, L and D as shown in the equivalent

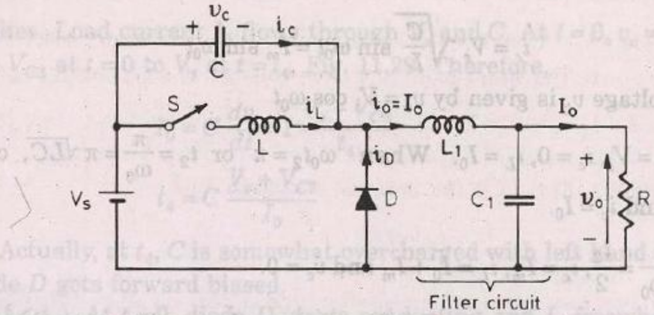


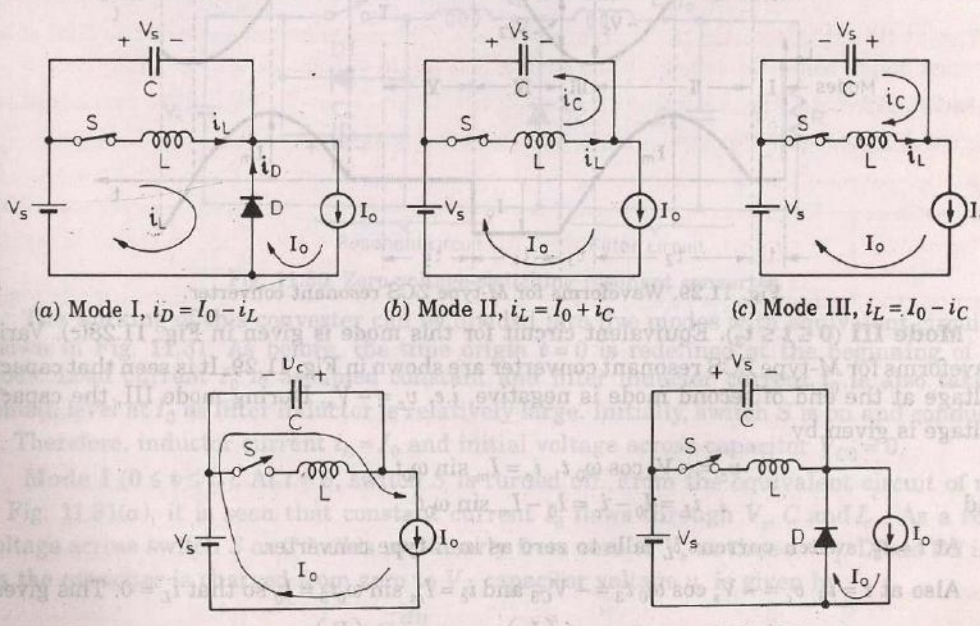
Fig. 11.27. *M*-type zero-current-switching resonant converter.

circuit of this mode, Fig. 11.28 (*a*). It is seen that $i_L = \frac{V_s}{L}t$ rises linearly as in *L*-type converter. Also, diode current $i_D = I_0 - i_L$

At $t = t_1$, $$i_L = I_0 = \frac{V_s}{L}t_1$$

or $$t_1 = \frac{I_0 \cdot L}{V_s} \qquad ...(11.21)$$

At $t = t_1$, $i_L = I_0$ and $i_D = I_0 - I_0 = 0$, diode *D* is therefore turned off. Voltage v_c across capacitor stays at V_s through *D*.



(*a*) Mode I, $i_D = I_0 - i_L$ (*b*) Mode II, $i_L = I_0 + i_C$ (*c*) Mode III, $i_L = I_0 - i_C$

(*d*) Mode IV, $v_c = V_{C3}$ at $t = 0$, $i_L = 0$ (*e*) Mode V, $i_L = 0$

Fig. 11.28. Equivalent circuits for *M*-type ZCS resonant converter.

Mode II ($0 \le t \le t_2$). After *D* turns off, load current I_0 flows through V_s and *L* as shown in Fig. 11.28(*b*). Also, *C* and *L* form a resonant circuit where the current i_c is given by

$$i_c = V_s \sqrt{\frac{C}{L}} \sin \omega_0 t = I_m \sin \omega_0 t$$

and capacitor voltage v_c is given by $v_c = V_s \cos \omega_0 t$

At $t = 0, v_c = V_s, i_c = 0, i_L = I_0$. When $\omega_0 t_2 = \pi$ or $t_2 = \frac{\pi}{\omega_0} = \pi\sqrt{LC}$, capacitor voltage $v_c = -V_s, i_c = 0$ and $i_c = I_0$.

When $t = \frac{\pi}{2\omega_0} = \frac{t_2}{2}, i_c = I_m, i_L = I_0 + I_m$ and $v_c = 0$.

During this mode, $i_L = I_0 + I_c = I_0 + I_m \sin \omega_0 t$ and capacitor gets charged from V_s to $-V_s$ as shown in Fig. 11.29.

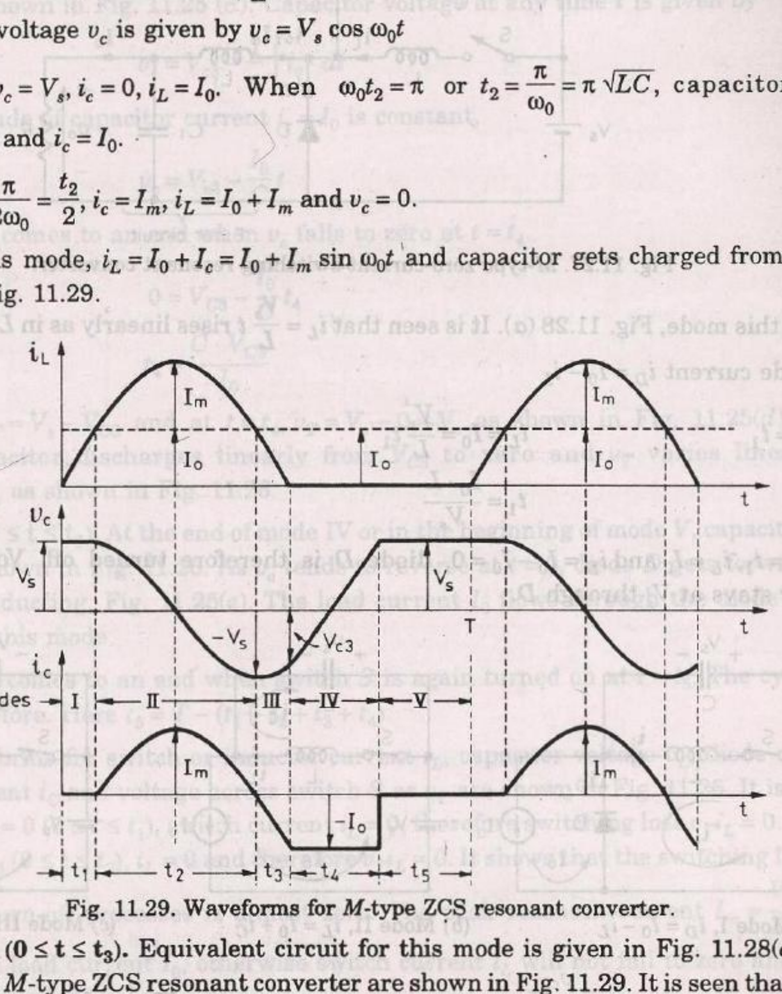


Fig. 11.29. Waveforms for *M*-type ZCS resonant converter.

Mode III ($0 \le t \le t_3$). Equivalent circuit for this mode is given in Fig. 11.28(*c*). Various waveforms for *M*-type ZCS resonant converter are shown in Fig. 11.29. It is seen that capacitor voltage at the end of second mode is negative, *i.e.* $v_c = -V_s$. During mode III, the capacitor voltage is given by

$$v_c = -V_s \cos \omega_0 t, \quad i_c = I_m \sin \omega_0 t$$

and
$$i_L = I_0 - i_c = I_0 - I_m \sin \omega_0 t$$

At $t = t_3$, switch current i_L falls to zero as in *L*-type converter.

Also at $t = t_3$, $v_c = -V_s \cos \omega_0 t_3 = -V_{C3}$ and $i_c = I_m \sin \omega_0 t_3 = I_0$ so that $i_L = 0$. This gives

$$t_3 = \frac{1}{\omega_0} \sin^{-1}\left(\frac{I_0}{I_m}\right) = \sqrt{LC} \sin^{-1}\left(\frac{I_0}{I_m}\right) \qquad ...(11.22)$$

Mode IV ($0 \le t \le t_4$). In the previous mode, as i_L falls to zero and tends to reverse, switch *S* is naturally turned off. In this mode, therefore, *S* remains off and the equivalent circuit of

Fig. 11.28(*d*) applies. Load current I_0 flows through V_s and C. At $t = 0$, $v_c = -V_{C3}$. Current I_0 charges C from $-V_{C3}$ at $t = 0$ to V_s at $t = t_4$, Fig. 11.29. Therefore,

$$I_0 = C\frac{dv}{dt} = C\frac{V_s + V_{C3}}{t_4}$$

or

$$t_4 = C\frac{V_s + V_{C3}}{I_0} \qquad \text{...(11.23)}$$

At t_4, $v_c = V_s$. Actually, at t_4, C is somewhat overcharged with left hand plate positive and consequently diode D gets forward biased.

Mode V ($0 \le t \le t_5$). At $t = 0$, diode D starts conducting and I_0 freewheels through D as shown in Fig. 11.28 (*e*). Switch S is open and voltage v_c stays at V_s through D. Switch current i_L remains zero as S is open. At time $t = T$, switch S is again turned on and the cycle repeats.

11.7.2. Zero-Voltage-Switching Resonant Converters

A zero-voltage-switching (ZVS) resonant converter is shown in Fig. 11.30. It consists of diode D1 and capacitor C connected across the switch S. As in ZCS converter, ZVS resonant converter has L, C as the resonant circuit components and L_1, C_1 as the filter circuit components. The function of resonant capacitor C is to produce zero voltage across the switch S. Diode D2 provides a free wheeling path to load current I_0. As the name suggests, the switch S in ZVS resonant converter is turned on and off at zero-voltage across the switch.

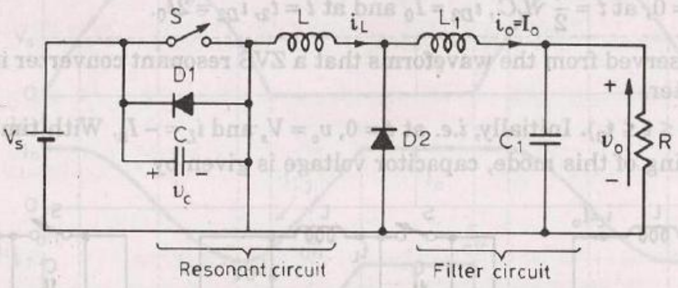


Fig. 11.30. Zero-voltage-switching resonant converter.

The working of this converter can be divided into five modes with equivalent circuits as shown in Fig. 11.31. As before, the time origin $t = 0$ is redefined at the beginning of each mode. Load current I_0 is assumed constant and filter inductor current i_0 is also taken to remain level at I_0 as filter inductor is relatively large. Initially, switch S is on and conducting I_0. Therefore, inductor current $i_L = I_0$ and initial voltage across capacitor $V_{C0} = 0$.

Mode I ($0 \le t \le t_1$). At $t = 0$, switch S is turned off. From the equivalent circuit of mode *I*, Fig. 11.31(*a*), it is seen that constant current I_0 flows through V_s, C and L. As a result, voltage across switch S *or* C builds up linearly from zero to V_s at time $t = t_1$. Diode D2 is off. As the capacitor is charged from zero to V_s, capacitor voltage v_c is given by

$$I_0 = C\frac{dv}{dt}$$

or

$$v_c = \frac{I_0}{C}t$$

At time $t = t_1$,

$$v_c = \frac{I_0}{C}t_1 = V_s \quad \text{or} \quad t_1 = \frac{CV_s}{I_0} \qquad \text{...(11.24)}$$

Note that voltage across diode D2 is $v_{D2} = V_s$ at $t = 0$ and $v_{D2} = 0$ at $t = t_1$.

Also, at $t = 0$, $v_c = 0$; therefore switch S is turned off at zero voltage as required.

Mode II ($0 \le t \le t_2$). At $t = 0$, actually capacitor is somewhat overcharged, *i.e.* $v_c > V_s$; therefore diode D2 becomes forward biased. Now a resonant current i_L is set up in series circuit V_s, C, L and D2, Fig. 11.31(*b*), where i_L is given by

$$i_L = I_0 \cos \omega_0 t$$

The capacitor voltage v_c is given by

$$v_c = V_s + V_m \sin \omega_0 t \qquad \text{...(11.25)}$$

where $V_m = I_0 \sqrt{\frac{L}{C}} = I_0 Z_0$ and $Z_0 = \sqrt{\frac{L}{C}}$ is the characteristic impedance of the circuit in ohms. The peak switch or capacitor voltage V_{pk} occurs when $\omega_0 t = \pi/2$ or $t = \frac{\pi}{2}\sqrt{LC}$ and its value is

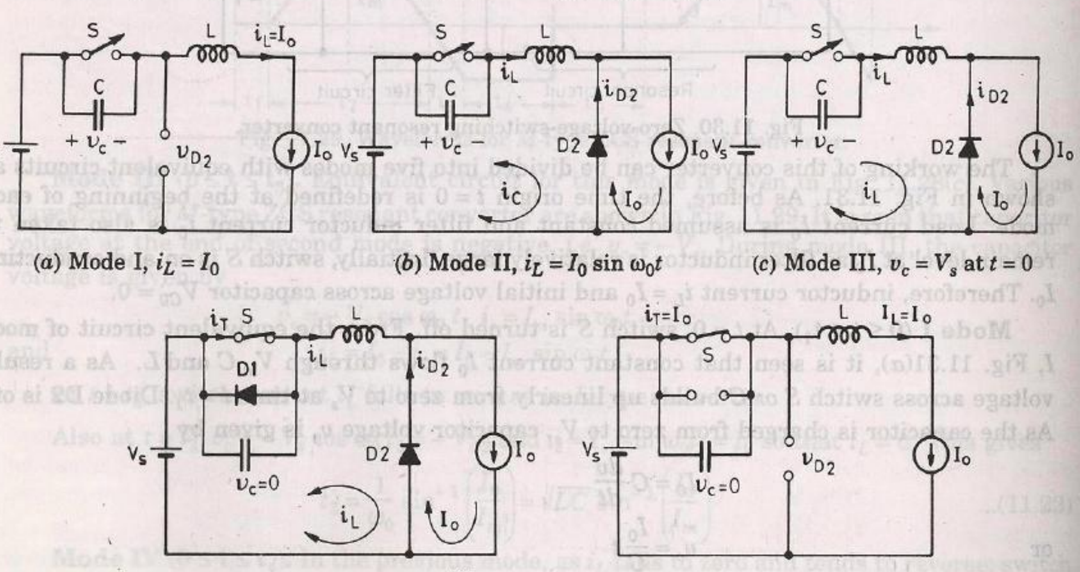
$$V_{pk} = V_s + V_m = V_s + I_0 Z_0$$

At $t = t_2$, $i_L = -I_0$ where $\omega_0 t_2 = \pi$ or $t_2 = \pi\sqrt{LC}$ and capacitor voltage is $v_c = V_s$. Diode D2 current is given by $i_{D2} = I_0 - I_0 \cos \omega_0 t$.

At $t = 0$, $i_{D2} = 0$, at $t = \frac{\pi}{2}\sqrt{LC}$, $i_{D2} = I_0$ and at $t = t_2$, $i_{D2} = 2I_0$.

It may be observed from the waveforms that a ZVS resonant converter is the dual of ZCS resonant converter.

Mode III ($0 \le t \le t_3$). Initially, *i.e.* at $t = 0$, $v_c = V_s$ and $i_L = -I_0$. With time t reckoned zero from the beginning of this mode, capacitor voltage is given by



(*a*) Mode I, $i_L = I_0$ (*b*) Mode II, $i_L = I_0 \sin \omega_0 t$ (*c*) Mode III, $v_c = V_s$ at $t = 0$

(*d*) Mode IV, $i_L = -I_{L3} + \frac{V_s}{L} t$ (*e*) Mode V, $i_T = i_L = I_0$

Fig. 11.31. Equivalent circuit for ZVS resonant converter.

$$v_c = V_s - V_m \sin \omega_0 t$$

and $$i_L = -I_0 \cos \omega_0 t$$

so that $$i_{D2} = I_0 - I_0 \cos \omega_0 t$$

At time $t = t_3$, $v_c = 0$, $i_L = I_{L3}$ and $i_{D2} = I_0 - I_{L3}$. This gives

$$0 = V_s - V_m \sin \omega_0 t_3$$

or $$t_3 = \sqrt{LC} \sin^{-1}\left(\frac{V_s}{I_0}\right)\sqrt{\frac{C}{L}} \quad ...(11.26)$$

At the end of this mode, *i.e.* at $t = t_3$, $v_c = 0$; as a result reverse bias across D1 vanishes and i_L begins to flow through D1.

Mode IV ($0 \le t \le t_4$). During this mode, capacitor voltage is clamped to zero by diode D1 conducting negative current i_L. As soon as antiparallel diode D1 begins to conduct at $t = 0$, gate drive is applied to switch S. The inductor current i_L rises linearly from $-I_{L3}$ to zero. At this instant, reverse bias of D1 vanishes and already gated switch S turns on. This shows that switch S turns on at zero voltage and zero current. After this, current rises linearly to I_0 in the circuit formed by V_s, S, L and D2. The linear variation of current from I_{L3} is given by

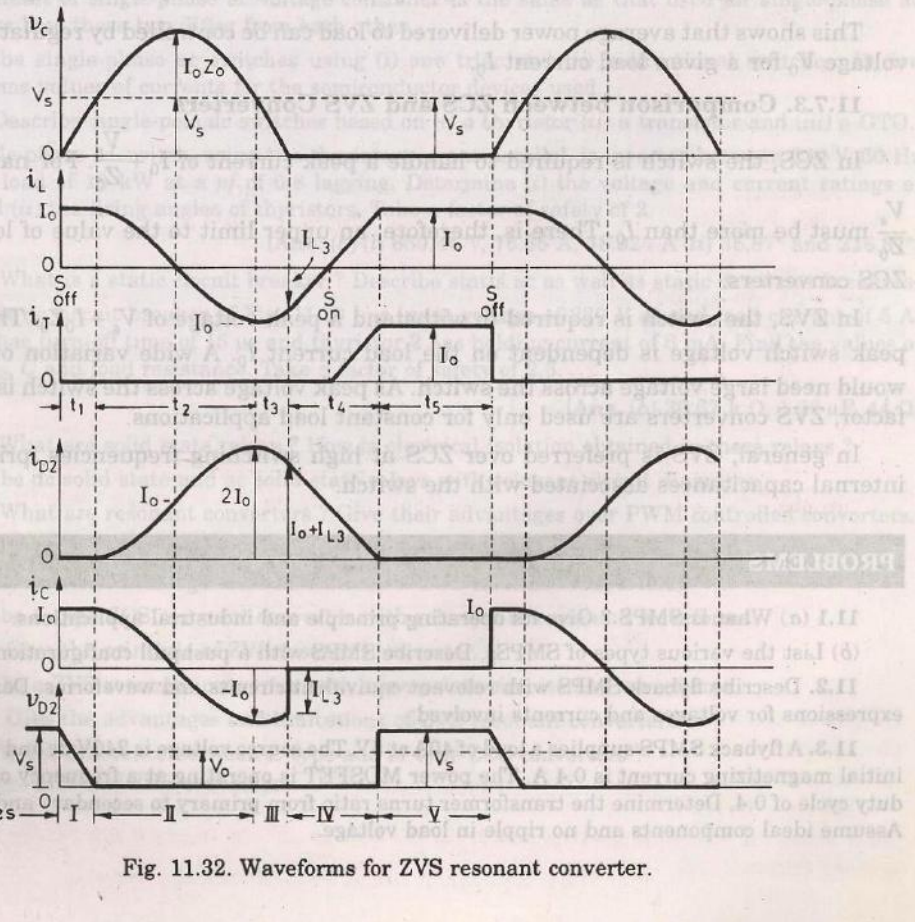


Fig. 11.32. Waveforms for ZVS resonant converter.

$$i_L = -I_{L3} + \frac{V_s}{L}t$$

At $t = t_4$, $i_L = I_0 = -I_{L3} + \frac{V_s}{L}t_4$. This gives

$$t_4 = (I_0 + I_{L3})\left(\frac{L}{V_s}\right) \quad ...(11.27)$$

Diode current $i_{D2} = I_0 + i_L$. At $t = 0$, $i_{D2} = I_0 + I_{L3}$ and time $t = t_4$, $i_{D2} = 0$. During modes II, III and IV, diode D2 is in conduction, therefore $v_{D2} = 0$, as shown in the waveforms of Fig. 11.32.

Mode V ($0 \le t \le t_5$). At the end of mode IV, or in the beginning of mode V at $t = 0$, i_L reaches I_0 and therefore diode D2 turns off. Switch S continues conducting I_0 as shown in Fig. 11.31(*e*). Note that voltage $v_{D2} = V_s$ during this mode. Mode V ends at $t = t_5$ when switch S is turned off again at zero voltage. The cycle now repeats as before.

The various waveforms for these five modes are now sketched in Fig. 11.32. It is seen from these waveforms that for a ZVS resonant converter :

(*i*) switch, or inductor, current is limited to I_0

(*ii*) average value of output voltage V_0 can be controlled by controlling the interval t_5.

This shows that average power delivered to load can be controlled by regulating the output voltage V_0 for a given load current I_0.

11.7.3. Comparison between ZCS and ZVS Converters

In ZCS, the switch is required to handle a peak current of $I_0 + \frac{V_s}{Z_0}$. For natural turn-off, $\frac{V_s}{Z_0}$ must be more than I_0. There is, therefore, an upper limit to the value of load current in ZCS converters.

In ZVS, the switch is required to withstand a peak voltage of $V_s + I_0Z_0$. This shows that peak switch voltage is dependent on the load current I_0. A wide variation of load current would need large voltage across the switch. As peak voltage across the switch is a dominating factor, ZVS converters are used only for constant load applications.

In general, ZVS is preferred over ZCS at high switching frequencies, primarily due to internal capacitances associated with the switch.

PROBLEMS

11.1 (*a*) What is SMPS ? Give its operating principle and industrial applications.

(*b*) List the various types of SMPSs. Describe SMPS with a pushpull configuration.

11.2. Describe flyback SMPS with relevant equivalent circuits and waveforms. Derive the various expressions for voltages and currents involved.

11.3. A flyback SMPS supplies a load of 40A at 5V. The source voltage is 240V dc and the transformer initial magnetizing current is 0.4 A. The power MOSFET is operating at a frequency of 50 kHz with a duty cycle of 0.4. Determine the transformer turns ratio from primary to secondary and its inductance. Assume ideal components and no ripple in load voltage.