$\therefore$ Battery kW = Battery voltage × Battery discharge current.

Now, total battery voltage = 10.6 × 12 = 127.2 V.

$$\therefore \quad \text{Battery discharge current, } I_{dc} = \frac{\text{Battery kW}}{127.2} = \frac{5.647 \times 10^3}{127.2} = 44.39 \text{ Amp.}$$

Capacity derating of battery up to 8 hr discharge is 0.5,

$$\therefore \quad \text{Back-up time} = \frac{\text{A-H capacity}}{\text{Discharge current}} = \frac{250 \text{ A-H}}{44.39} = 4.864 \text{ hours.}$$

(ii) *Charger Peak Rating*

Assume that the battery is completely discharged during the back-up time of 4.864 hrs.

$\therefore$ AH output from the battery = 250 A-H

Now, in order to charge the battery completely,

A-H input = A-H output

i.e. charging current × No. of charging hours = 250.

$$\therefore \quad \text{charging current} = \frac{250}{4} = 62.50 \text{ Amp.}$$

$\therefore$ Charger peak power = 144 × 62.5 = 9 kW.

16.3 SWITCHED MODE POWER SUPPLIES (SMPS)

In the low to medium power range, say, upwards from 50 W, a d.c. source of supply is often required which contains negligible a.c. ripple which can be controlled in magnitude. For this application, switched mode power supplies circuits are used. Though the SMPS was developed originally by NASA in the 1960s to provide a compact power supply system for space vehicles, it had no significant impact on the general power supply market until the 1970s. From that point development was rapid and the SMPS now accounts for some 75% of the power supplies being produced.

An SMPS is based on a d.c. chopper with a rectified and possibly transformed output. The output voltage amplitude is controlled by varying the mark–space ratio of the chopper. This may be achieved by means of pulse-width control or frequency variation with constant pulse width, the former being the more common. The circuit techniques used for SMPS can be separated into following four broad categories.

(i) Flyback SMPS
(ii) Feed-forward SMPS
(iii) Push-pull SMPS
(iv) Bridge SMPS

Figure 16.6 Shows the building blocks of a typical high-frequency off-the line switching power-supply. As the name implies, the input rectification in a switching power supply is done directly off-the line, without the use of the low-frequency isolation power transformer, as in the case of linear power supply. The a.c line is directly rectified and filtered to produce a raw high-voltage d.c., which in turn is

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fed into a switching element. The switch is operating at the high-frequencies of 20 kHz to 1 MHz, chopping the dc voltage into a high-frequency square-wave. This square-wave is fed into the power isolation transformer, stopped down to a predetermined value and then rectified and filtered to produce the required dc output.

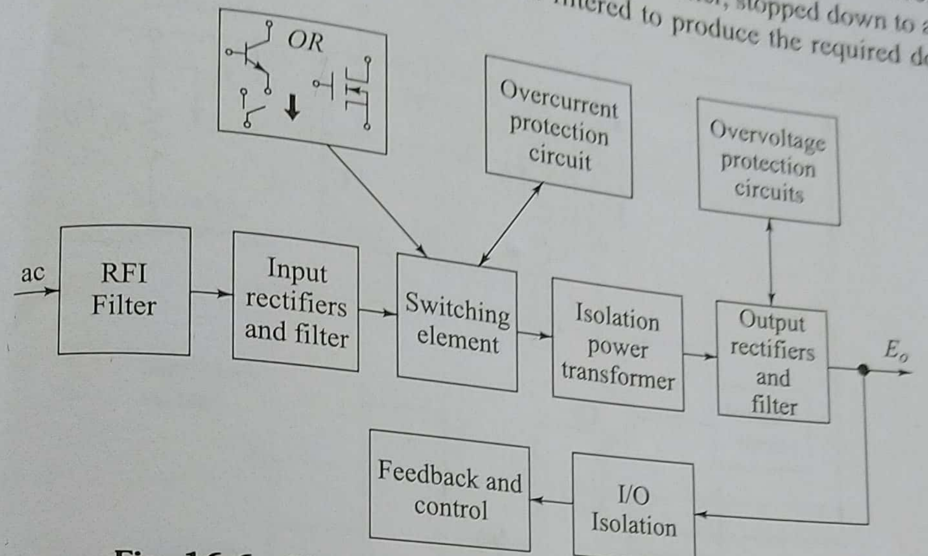


Fig. 16.6 *Block diagram of off-line high frequency SMPS*

A portion of this output is monitored and compared against the fixed reference voltage, and the error signal is used to control the on–off times of the switch, thus regulating the output. Since the switch is either ON or OFF, it is dissipating very little energy, resulting in a very high overall power supply efficiency of about 70 to 80%. Another advantage is the power-transformer size which can be quite small due to high operating frequency. Hence the combination of high efficiency (i.e. no large heat sinks) and relatively small magnetics, results in compact, light weight power supplies, with power densities up to 30 W/in^3 versus 0.3 W/in^3 for linears.

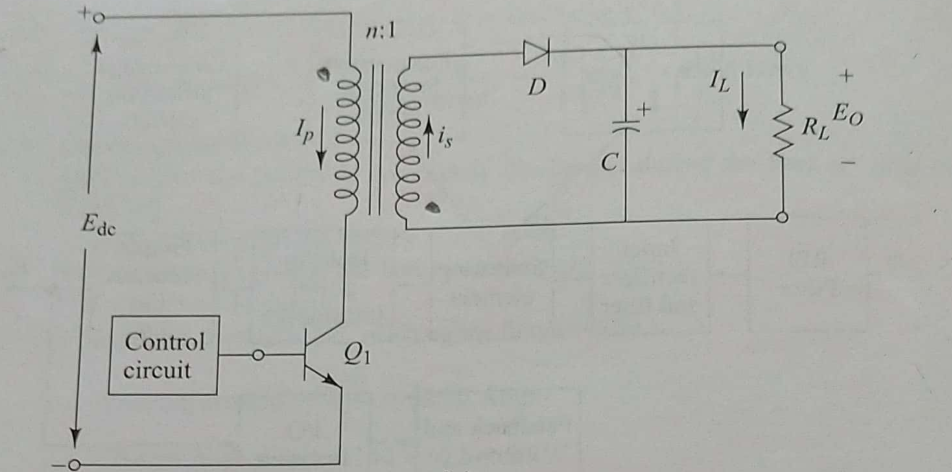
16.3.1 Isolated Flyback Converters

Although there are numerous converter circuits described by number of authors and researchers, basically all of them are related to three classical circuits known as the "flyback or buck-boost", the "forward or buck", and the "push-pull or buck derived" converter. The buck-boost regulators (Chapter 8) comes under the category of nonisolated SMPS. In these nonisolated regulators, the isolation between the load and source is provided only by the power-semiconductor devices, therefore if these devices fail then the load will be directly connected across the source and may damage it completely.

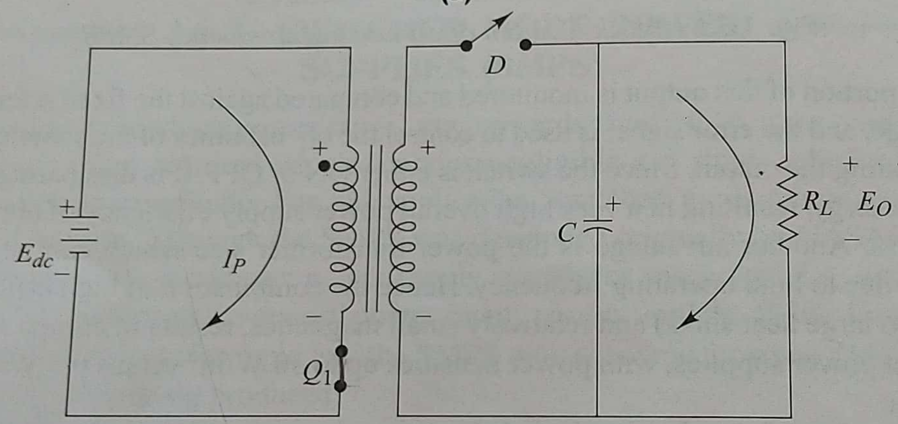
In isolated converters, the ferrite core high-frequency transformer is used for electrical isolation between the load and the source. Another advantage of using a transformer is the possibility to get multiple outputs.

16.3.1.1 Discontinuous Mode (Flyback Converter)

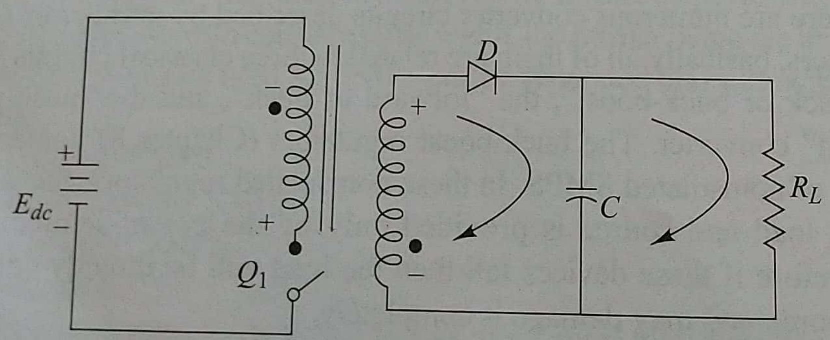
The circuit diagram for the discontinuous flyback converter is shown in Fig. 16.7 and associated steady-state waveforms are shown in Fig. 16.8. The circuit operates as follows:



(a)



(b) Mode I equivalent circuit



(c) Mode II equivalent circuit

Fig 16.7 *Flyback converters*

erter)
verter is shown in Fig.
n Fig. 16.8. The circuit

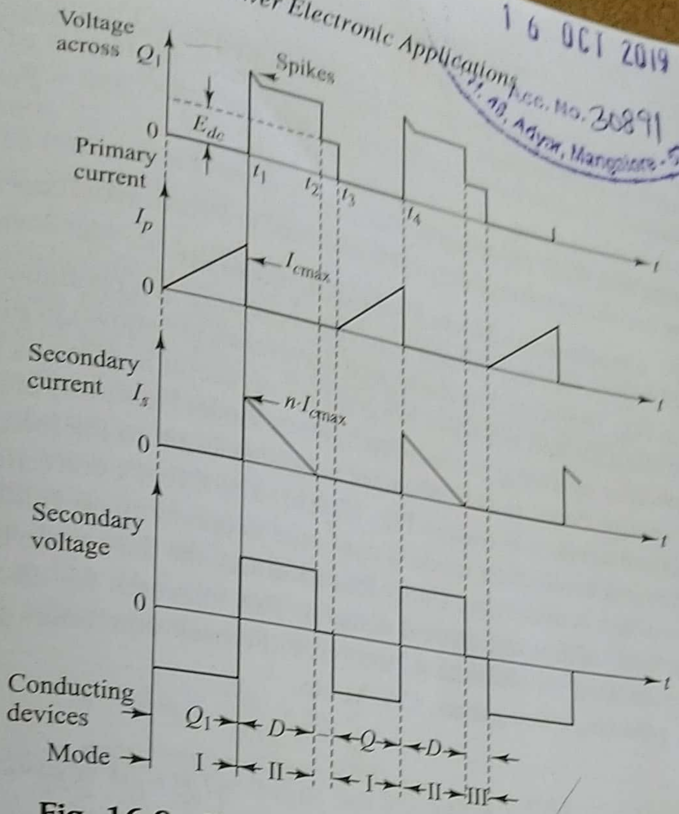


Fig. 16.8 *Flyback converter associated waveforms.*

Mode I ($t_0 - t_1$): When transistor Q_1 is ON at $t = 0$, primary current starts to build-up in the primary winding, storing energy. Due to the opposite primary arrangement between the input and output windings of the transformer choke, there is no energy transferred to the load since diode D is reverse-biased. Transistor Q_1 is turned-off abruptly at instant $t = t_1$ The equivalent circuit for this mode is shown in Fig. 16.7 (b).

Mode II ($t_1 - t_2$): When the transistor is turned-off at $t = t_1$, the polarity of the windings reverses due to the collapsing magnetic field. Now, diode D is conducting, charging the output capacitor C and delivering current I_L to load.

The voltage across Q_1 is the sum of input supply voltage (E_{dc}) and the self induced voltage across the primary winding (Ldi/dt), therefore it is higher than the supply voltage (E_{dc}). The equivalent circuit for this mode is shown in Fig. 16.7(c). The secondary current goes to zero at $t = t_2$. Thus, the stored energy in the transformer core is delivered to the load during this mode of operation.

(iii) Mode III ($t_2 - t_3$): In this mode, transistor and diode both are in the off-state. Therefore, primary and secondary currents are zero. As there is no voltage drop across the primary winding of the transformer, the voltage across the transistor Q_1 is equal to the dc supply voltage (E_{dc}). The secondary voltage is zero. The one-cycle operation completes in this mode and repeats itself.

Advantages of Discontinuous Mode

(i) Slower diodes can be used on the secondary side for rectification.

(ii) Size of transformer is smaller than that in the continuous mode.

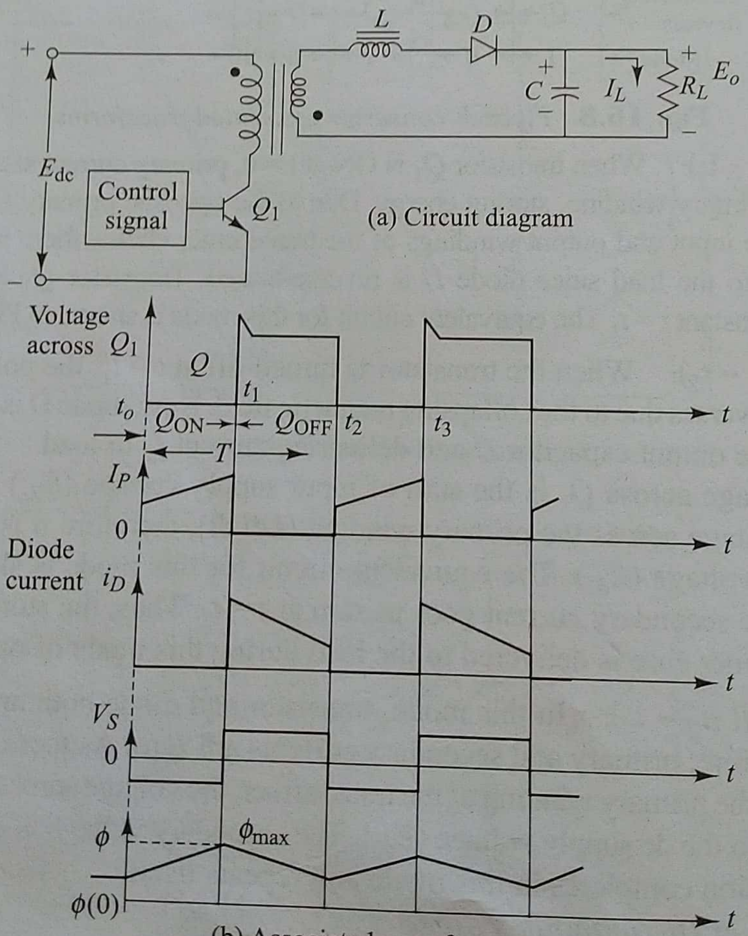
(iii) An additional filter inductance is not required on the secondary side.
(iv) Fast transient response.

Disadvantages

(i) Peak current rating of both devices is high.
(ii) Large size filter capacitor is required.
(iii) The maximum value of duty cycle is limited to 50% to avoid core saturation.

16.3.1.2 Continuous Mode Flyback Converter

The circuit diagram for the flyback converter operating in a continuous mode is as shown in Fig. 16.9(a) and its associated waveforms are shown in Fig. 16.9(b). The circuit diagram and the basic principle of operation is same as that of the flyback converter operating in the discontinuous mode. However, in continuous mode, the energy stored in core is completely transferred to the load, transistor Q1 is turned-on again as shown in Fig. 16.9(b). The flyback converter operates in the continuous conduction mode if converter is operating on full-load and if the supply voltage is minimum. Under this condition, the duty cycle is increased to meet the load voltage and current demand. This increases the on-time of the pulse and decreases the off-time of the pulse. Flyback converters are used in computers, printers, video games, CATV, etc.



(a) Circuit diagram

(b) Associated waveforms

Fig. 16.9 *Flyback converter continuous mode.*

(i) Advantages of Continuous Mode:

(i) The size of the filter capacitor required for this mode in the secondary side is nearly half the size of the filter capacitor used in discontinuous mode.

(ii) The peak current rating of the diode and the transistor reduces to 50% of that used in the discontinuous mode.

(ii) Disadvantages of Continuous Mode:

(i) Rectifier diodes should be nearly four times faster than discontinuous current mode (typical $t_{rr} = 25–100$ ns).

(ii) Size of the transformer is larger than that of the discontinuous mode.

(iii) Expression for Voltage Transfer Ratio:

Let, E_{dc} be the dc input voltage to the converter,
N_1/N_2 be the transformer turns ratio,
E_0 be the output voltage and
α be the duty ratio of the transistor.

As shown in Fig. 16.9, the inductor core flux increases linearly from its initial value $\phi(0)$, which is finite and positive and is given by

$$\phi_{(t)} = \phi_{(0)} + \frac{E_{dc}}{N_1}\, t, \text{ for } 0 < t < t_{on} \tag{16.6}$$

The value of peak flux at the end of this interval is given by

$$\phi_{max} = \phi(t_{on}) = \phi_{(0)} + \frac{E_{dc}}{N_1}\, t_{on} \tag{16.7}$$

Transistor is turned-off after t_{on} and the energy stored in the transformer core causes the current to flow in the secondary winding. The voltage across the secondary $e_2 = -E_0$, and therefore the flux decreases linearly during the off-time of switch, i.e. t_{off}.

$$\therefore \qquad \phi_{(t)} = \phi_{max} - \frac{E_o}{N_2}(t - t_{on}) \text{ for } t_{on}<t<T \tag{16.8}$$

and
$$\phi_{(T)} = \phi_{max} - \frac{E_o}{N_2}(T - t_{on}) \tag{16.9}$$

Substituting ϕ_{max} from (16.7) in (16.9), we get

$$\phi_{(T)} = \phi_{(0)} + \frac{E_{dc}}{N_1}\, t_{on} - \frac{E_o}{N_2}(T - t_{on}) \tag{16.10}$$

In steady state, since the net change of flux through the core over one time period must be zero,

$\phi_{(T)} = \phi_{(0)}$, substitute in Eq. (16.10),

$$\therefore \qquad \phi_{(0)} = \phi_{(0)} + \frac{E_{dc}}{N_1}\, t_{on} - \frac{E_o}{N_2}(T - t_{on})$$

$$\frac{E_{dc}}{N_1} t_{on} = \frac{E_o}{N_2} (T - t_{on})$$

$$\therefore \quad \frac{E_o}{E_{dc}} = \frac{t_{on}}{(T - t_{on})} \frac{N_2}{N_1} = \frac{t_{on}/T}{(1 - t_{on}/T)} \frac{N_2}{N_1}$$

but, $\alpha = t_{on}/T,$

$$\therefore \quad \frac{E_o}{E_{dc}} = \frac{\alpha}{1 - \alpha} \frac{N_2}{N_1} \tag{16.11}$$

Equation (16.11) shows that the voltage transfer ratio in flyback converter depends on α in an identical manner as in the buck-boost converter.

(iv) Selection Criteria for Switching Transistor:

(a) The switching transistor used in the flyback converter must be chosen to handle peak collector voltage at turn-off and peak collector currents at turn-on. The peak collector voltage which the transistor must sustain at turn-off is

$$V_{CE,max} = \frac{E_{dc}}{1 - \alpha_{max}} \tag{16.12}$$

From Eq. (16.12), it becomes clear that in order to limit the collector voltage to a safe value, the duty cycle must be kept relatively low, normally below 50 per cent, i.e. $\alpha_{max} < 0.5$.
In practice, α_{max} is taken at about 0.4, which limits the peak collector voltage $V_{CE,max} < 2.2\, E_{dc}$ and, therefore, transistors with working voltages above 800 V are usually used in the off-line flyback converter designs.

(b) The Selected transistor must have,

$$I_c = \frac{I_L}{n} = I_p \tag{16.13}$$

where I_p is the primary transformer-choke peak current,
n is the primary to secondary turns ratio,
and I_L is the output load current.

(v) Selection of Transformer-Choke:

Since the transformer-choke of the flyback converter is driven in one direction only of the B-H characteristics curve, it has to be designed so that it will not saturate. Therefore, a core with a relatively large volume and air-gap must be used. The effective transformer-choke volume is given by

$$\text{Volume } (V) = \frac{\mu_o \cdot \mu_e \cdot I_{L,max}^2 \cdot L_{out}}{\beta_{max}^2} \tag{16.14}$$

where, $I_{L,max}$ = determined by load current.
μ_e = relative permeability of the chosen core material.
β_{max} = maximum flux density of the core.

The relative permeability μ_e must be chosen to be large enough to avoid excessive temperature rise in core due to restricting core and wire size and therefore copper and core-losses.

(vi) Expression for Peak Working Collector Current:

The energy transferred in the choke is given by

$$P_{out} = \left(\frac{L\, I_p^2}{2T}\right) n \tag{16.15}$$

where n = efficiency of the converter.

The voltage across the transformer may be expressed as

$$E_{dc} = L\,\frac{di}{dt} \tag{16.16}$$

Let us take $\quad di = I_p$ and $\dfrac{1}{dt} = \dfrac{f}{\alpha_{max}}$

Substitute in Eq. (16.16), we get

$$E_{dc} = \frac{L\, I_p \cdot f}{\alpha_{max}} \tag{16.17}$$

or

$$L = \frac{E_{dc} \cdot \alpha_{max}}{I_p \cdot f} \tag{16.18}$$

Substituting Eq. (16.18) into Eq. (16.15), we get

$$P_o = \left(\frac{E_{dc} \cdot f \cdot \alpha_{max} \cdot I_p^2}{2 \cdot f \cdot I_L}\right) \eta = \frac{1}{2} \eta\, E_{dc}\, \alpha_{max}\, I_p$$

or,

$$I_p = \frac{2\, p_o}{\eta\, E_{dc} \cdot \alpha_{max}} \tag{16.19}$$

Substituting Eq. (16.19), in Eq. (16.13), we get

$$I_c = \frac{2\, p_o}{\eta\, E_{dc} \cdot \alpha_{max}} \tag{16.20}$$

For converter with efficiency 80% and duty-cycle

$\alpha_{max} = 0.4$, we get

$$I_c = \frac{6.2 \cdot P_o}{E_{dc}} \tag{16.21}$$

16.3.1.3 *Variations of Basic Flyback Converter*

We have seen that with the basic flyback circuit, the collector voltage of the switching transistor must sustain at least twice the input voltage at turn-off. In cases where the voltage value is too high to use a commercial transistor type, the two-transistor flyback converter may be used, as shown in Fig. 16.10. This circuit uses two transistors, which are switched ON or OFF simultaneously. Diodes D_1 and D_2 act as clamping diodes restricting the maximum collector voltage of the transistor to E_{dc}. Thus, lower voltage transistors may be used to realize this design, but at the expense of three-extra components, i.e. Q_2, D_1 and D_2.

An advantage of the flyback circuit is the simplicity by which a multiple output switching power supply may be realized. This is because the isolation element acts as a common choke to all outputs, this only a diode and a capacitor are needed for an extra output voltage. Figure 16.11 illustrates a practical circuit.

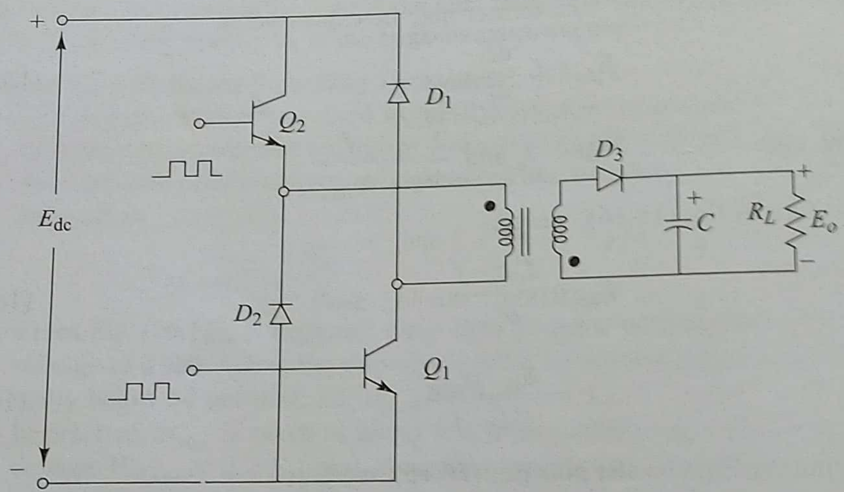


Fig. 16.10 *Two transistor flyback circuit*

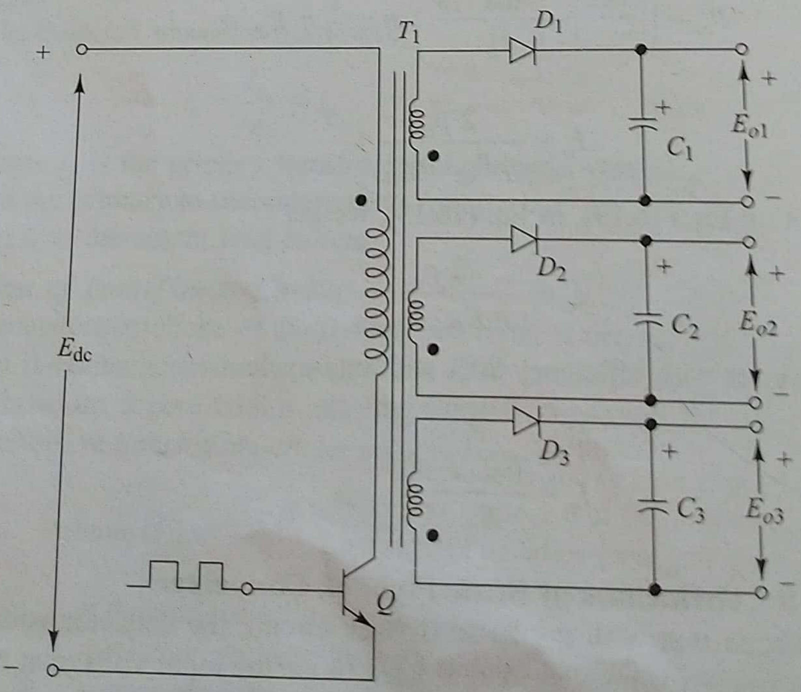


Fig. 16.11 *Flyback converter with multiple outputs*

SOLVED EXAMPLES

Example 16.6 An isolated flyback converter operating in continuous conduction mode at a frequency of 60 kHz is fed from the supply mains via a full-wave bridge rectifier-cum-capacitor filter and uses a 1200 V power-MOSFET as switching element. Assuming a ripple-free output of capacitor filter and neglecting rectifier drops, determine:

(i) Maximum duty-cycle, allowing a margin of 180 V for voltage spikes on the drain of power MOSFET for a nominal mains voltage of 230 V, 50 Hz.

(ii) The flyback converter turns ratio required to obtain a maximum output voltage of 12 V for a minimum mains voltage of 180 V.

Solution: *Given:* Switching frequency, $f_s = 60$ kHz., E_{sp} = Spike voltage = 180 V.

Mains = 230 V, 50 Hz, $E_0 = 12$ V

(i) *Calculation of α_{max}:*

From Eq, 16.12, voltage across switch is given by

$$E_{DS} = \frac{E_{dc}}{1-\alpha_{max}} + \text{spike voltage}$$

Voltage across the MOSFET should not exceed 1200 V.

$\therefore \quad E_{DS} < 1200$ Volts.

$$\therefore \quad E_{DS} = \frac{E_{dc}}{1-\alpha_{max}} + E_{sp} < 1200 \text{ V}$$

E_{dc} at 230V mains voltage with capacitor filter $= 230 \times \sqrt{2}$

$$\therefore \quad \frac{230 \times \sqrt{2}}{1-\alpha_{max}} + 180 < 1200 \text{ V}, \quad \frac{325.26}{1-\alpha_{max}} < 1020$$

$$\therefore \quad \alpha_{max} = 0.68$$

(ii) *Calculation of turns-ratio (N_1/N_2):*

$$\text{From Eq. 16.11, } E_0 = \frac{\alpha}{1-\alpha}\left(\frac{N_2}{N_1}\right)E_{dc}$$

$$\therefore \quad \frac{N_1}{N_2} = \frac{\alpha}{1-\alpha}\frac{E_{dc}}{E_o} = \left(\frac{0.68}{1-0.68}\right)\left(\frac{230 \times \sqrt{2}}{12}\right) = 57.6.$$

Example 16.7 A flyback converter is operated in a continuous mode from a supply of 14V to 30V with two outputs 12 V at 0.6A and –12 V at 0.6 A. The switching power supply is used to power some drivers that have intermittent load demands. The load can vary from 0.1 to 0.5 Amp. Assuming the efficiency of the converter to be 80% and switching frequency to be 50 kHz, determine:

(i) The average input power and current. (ii) Ratings of the transistor.

(iii) The primary widing inductance and the number of turns of the primary and the secondary if the core exhibits 80 MH per 1100 turns.

(iv) Ratings of the rectifying diode.

Solution:

(i) Average input power and current:

The full-load input power is given by

$P_o = (12 \times 0.6) + (12 \times 0.6) = 14.4$ W.

$$\text{Input power } P_i = \frac{P_o}{\eta} = \frac{14.4}{0.8} = 18 \text{ W}$$

Now, maximum value of average input current,

$$I_{(avg)_{max}} = \frac{P_i}{E_{dc_{min}}} = \frac{18}{14} = 1.285 \text{ Amp.}$$

Similarly, minimum value of average input current,

$$I_{i(av)_{min}} = \frac{P_i}{E_{dc_{max}}} = \frac{18}{30} = 0.6 \text{ A}$$

(ii) *Ratings of transistor:*

$$\text{From Eq. 16.20, } I_{c(max)} = \frac{2P_o}{\eta\, E_{dc_{min}} \cdot \alpha_{max}} \quad \text{Assume, } \alpha_{max} = 0.5$$

$$I_{c_{max}} = \frac{2 \times 14.4}{0.8 \times 14 \times 0.5} = 5.14 \text{ Amp.}$$

$$\text{Now,} \quad V_{CE_{max}} = \frac{E_{dc_{max}}}{1-\alpha_{max}} = \frac{30}{1-0.5} = 60 \text{ V}$$

$$P_{d_{max}} = V_{CE_{max}} \times I_{c_{max}} = 60 \times 5.14 = 308 \text{ W.}$$

(iii) Primary winding inductance and turns:

The value of primary inductance is given by

$$L_{prim} = \frac{E_{dc_{(min)}}}{I_{c_{max}} \cdot f_s}\, \alpha_{max} \qquad (16.22)$$

$$= \frac{14}{5.14 \times 50 \times 10^3} \times 0.5 = 27\ \mu\text{H.}$$

Since the core exhibits 80 MH per 1100 turns,

$$\therefore \quad N_{prim(N1)} = 1100\sqrt{\frac{L_{prim.}}{L_{1100}}} = 1100\sqrt{\frac{27 \times 10^{-6}}{80 \times 10^{-3}}}$$

$$= 21 \text{ turns.}$$

From Eq. (16.11),

$$N_2 = 18$$

(iv) *Diode ratings:*

$$\text{Peak diode current } I_D \geq \left(\frac{N_1}{N_2}\right) \cdot I_p \geq \left(\frac{N_1}{N_2}\right) \cdot I_{c_{max}}$$

$$\geq \left(\frac{21}{18}\right) \cdot (5.14) \geq 5.6 \text{ Amp}$$

$$\text{PIV} \geq E_0 + \frac{E_{dc(max)}}{\eta} \geq 12 + \frac{30}{(21/18)} \geq 37.71 \text{ V} \approx 40 \text{ V}$$

(16.22)

16.3.2 Isolated Forward Converter

Figure 16.12(a) shows the basic circuit diagram of a single ended isolated forward converter. At first glance, this power circuit resembles that of the flyback converter. However there are some distinct differences between the two circuits:

(i) The dot polarities are on the same side of the transformer. This means that the two windings are wound in the same sense and hence carry the current simultaneously. Thus, the transformer of the forward converter is a pure transformer.

(ii) Since the transformer is merely a transformer, this suggests a need of a filter inductance on the secondary side as shown in Fig. 16.12(a). The circuit operates as follows:

(i) Mode I (Q_1 ON) As soon as Q_1 is turned-on, the supply voltage E_{dc} is applied across the primary winding of the transformer. Due to this constant voltage, the primary current increases at a constant rate. Due to the winding polarity as shown in Fig. 16.12(a), the induced voltage in the secondary winding will forward bias diode D_1 and the secondary current starts flowing.

(ii) Mode II (Q_1 OFF) When Q_1 is turned-off, the primary voltage will change its polarity as shown in Fig. 16.12(b). The secondary voltage also will change its polarity. Diode D_1 is reversed-biased and D_2 is forward biased due to the induced voltage in the filter inductance and the current flows through the load as shown in Fig. 16.12(c).

Disadvantage: With basic configuration of Fig. 16.12(a), some residual energy remains in the transformer core. Due to this residual energy, the transformer core will saturate after a few cycles of operation. This will lead to overcurrent through Q_1 and may damage Q_1. The core saturation can be avoided only by removing the residual energy from the core. This is achieved by adding a "tertiary" or "demagnetizing" winding into the basic configuration of forward-converter.

16.3.2.1 Forward Converter with Tertiary Winding Figure 16.13(a) shows the circuit diagram for the forward converter with tertiary winding added to the basic configuration. Operation of the circuit is same as the basic forward converter, except for the operation of the demagnetizing winding. Such converters are used in computers, wordprocessor, televisions, etc.

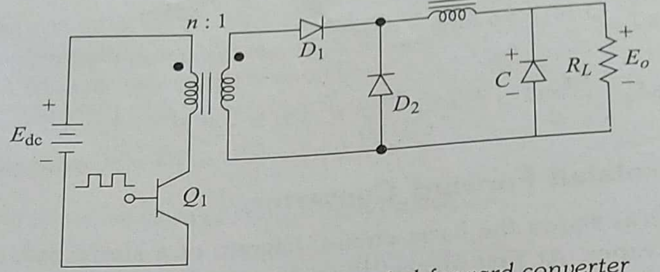


Fig. 16.12(a) *Basic single-ended forward converter*

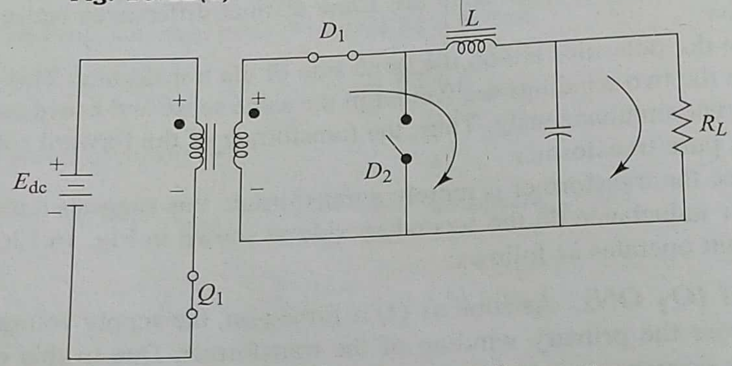


Fig. 16.12(b) *Mode I (Q1—ON) operation*

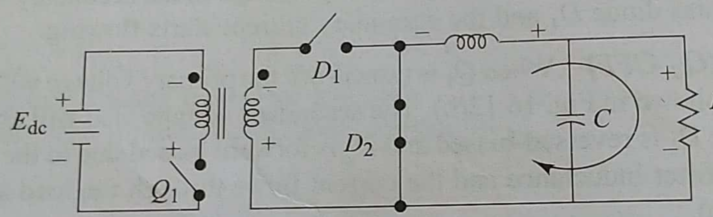


Fig. 16.12(c) *Mode-II (Q_1—OFF) operation*

Operation of Tertiary Winding: When the transistor Q_1 is turned-on, due to the winding polarities, diode D_m is reverse-biased and does not conduct. However, when Q_1 is turned-OFF, D_m is forward-biased and the current flows through the tertiary winding as shown in Fig. 16.13(b). The residual energy in the transformer core is returned back to the dc source via diode D_m and the tertiary winding. This is how the tertiary winding helps to demagnetize the core and avoids the core saturation. Due to the dot convention as shown in Fig. 16.13(a), the primary winding and tertiary winding will never carry current simultaneously. Associated waveforms are shown in Fig. 16.14. The dark-areas on the waveforms of Fig. 16.14 shows the magnetizing-demagnetizing current, given as,

$$I_{mag} = \frac{t_{on} \cdot E_{dc}}{L} \tag{16.23}$$

where L is the output inductance in microhenries.

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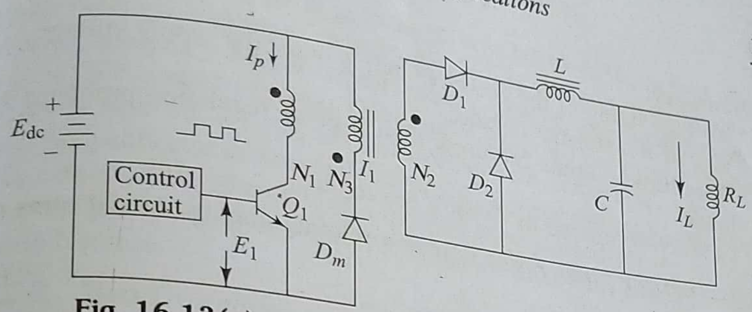

Fig. 16.13(*a*) *Forward converter with tertiary winding*

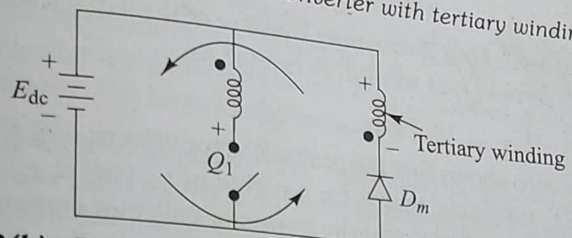

Fig. 16.13(*b*) *Demagnetization of transformer core by tertiary winding*

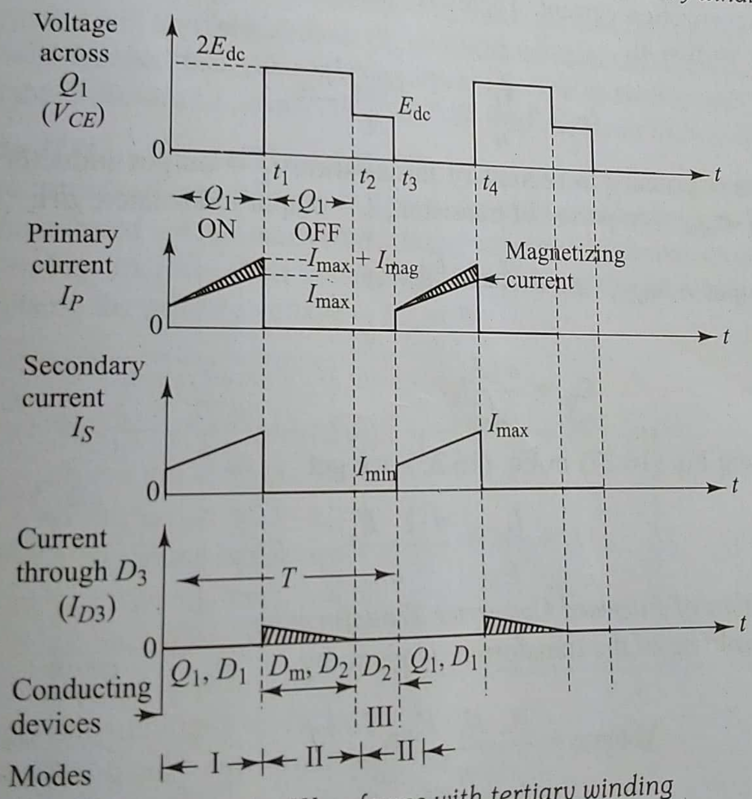

Fig. 16.14 *Waveforms with tertiary winding*

Advantages:

(i) It needs only one power transistor as switching element.

(ii) Simple driver circuits are required as compared to forward converters.

Disadvantages:

(i) Voltage rating of the power switch is twice the supply voltage E_{dc}.
(ii) Size of transformer is large.
(iii) If multiple outputs are used, then it gives regulation problems at light loads.
(iv) Costlier compared to flyback converters.

(i) Expression for turns ratio: Equation relating duty cycle and turns ratio is given by

$$\alpha_{max} = \frac{1}{1 + N_3/N_1} \tag{16.24}$$

(ii) Selection of Switching Transistor: Because of the tertiary winding and diode D_3, the voltage across Q_1 at turn-off is limited to

$$V_{CE,\,max} = 2\,E_{dc} \tag{16.25}$$

The waveforms also shows that the peak collector voltage of 2 E_{dc} is maintained for as long as D_m conducts, that is for a period of t_{on} $(=T\cdot\alpha_{max})$. Also, it can be observed from the waveforms that the transistor collector current at turn-on will have a value equal to that derived for flyback converter plus the net amount of the magnetization current. Therefore, peak collector current in the transistor may be written as,

$$I_{C_{max}} = \frac{I_L}{n} + \frac{T\cdot\alpha_{max}\cdot E_{dc}}{L} \tag{16.26}$$

where, n = primary to secondary turns ratio., I_L = output inductor current, Amp., $T\cdot\alpha_{max}$ = on-period of transistor., L = output inductance, μH.

Also, output voltage, $$E_0 = \frac{\alpha_{max}\cdot E_{dc}}{n}$$

or, $$E_{dc} = \frac{n\cdot E_o}{\alpha_{max}} \tag{16.27}$$

Substituting Eq. (16.27) in Eq. (16.25), we get

$$I_c = \frac{I_L}{n} + \frac{n\cdot T\cdot E_o}{L} \tag{16.28}$$

(iii) Selection of Forward Converter Transformer:

The core volume of the transformer is given by

$$\text{Volume} = \frac{\mu_o\cdot\mu_e\cdot I^2_{mag.}\,L}{B^2_{max}} \tag{16.29}$$

where, $$I_{mag} = \frac{n\cdot T\cdot E_o}{L} \tag{16.30}$$

It should be noted that the α_{max} must be kept below 50%, so that when the transformer voltage is clamped through the tertiary winding, the integral of the volt-seconds between the input voltage, when Q_1 is ON, and the clamping level, when Q_1 is OFF, amounts to zero. Duty cycles above 50 per cent, will upset the volt-seconds balance, driving the transformer into saturation, which in turn produces high collector current spikes that may destroy the switching transistor.

Also, care must be taken during construction to couple the tertiary winding tightly to the primary (bifilar wound) to eliminate fatal voltage spikes caused by leakage inductance.

16.3.2.2 Two-Transistor Forward Converters

Figure 16.15(a) shows the circuit diagram of the two-transistor forward converter, and the associated waveforms are shown in Fig. 16.15(b). This configuration reduces the voltage ratings of the transistor to $E_{dc(max)}$ instead of 2 $E_{dc(max)}$ (Single-ended configuration). The circuits operate as follows:

(i) Mode I (Q_1, Q_2 and D_3 ON): At $t = 0$, transistors Q_1 and Q_2 are turned on simultaneously. The supply voltage E_{dc} is connected across the primary winding. The primary current starts increasing linearly from I_{min} to $I_{max} + I_{mg}$, where I_{mg} is the magnetizing component shown by shaded area in the primary current waveform. Due to the specific winding directions, the induced voltage in the tertiary winding will reverse bias diode D_m and the induced voltage in the secondary winding will forward bias the rectifying diode D_3. The secondary will deliver power to the inductance L, capacitor C and the load as shown in Fig. 16.15(b).

(ii) Mode II (Q_1, Q_2, OFF and D_4. D_m): At $t = t_1$, both the transistors Q_1 and Q_2 are turned-off simultaneously. Due to the sudden interruption of primary current, the induced voltage across the primary winding will change its polarities as shown in Fig. 16.15(b). This voltage will forward bias the diodes D_1, D_2 and they will clamp the primary voltage to E_{dc} volts.

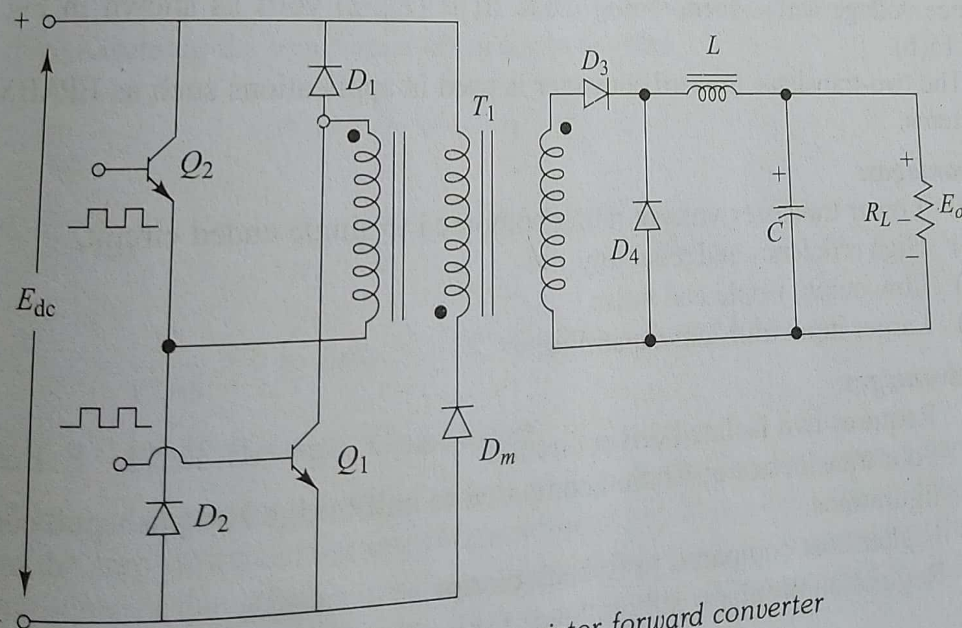

Fig. 16.15(a) *Two-transistor forward converter*

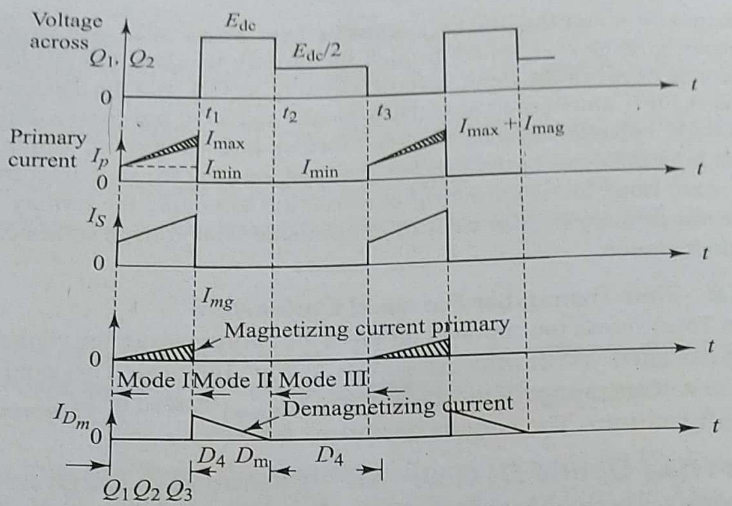

Fig. 16.15*(b)* *Voltage and current waveforms*

The induced voltages across the secondary and tertiary windings will also change their polarities. This will forward bias diode D_m and reverse bias D_3. The demagnetizing current starts flowing through the tertiary winding and D_m as shown in Fig. 16.15(b). Due to the induced voltage across the filter inductance L, diode D_4 is forward biased and the load current is maintained by the inductance L and filter capacitor C as shown in Fig. 16.15(b).

(iii) Mode III (D_4 ON): At $t = t_2$, the demagnetizing current reduces to zero so diode D_m is turned-off. All the other devices except D_4 are in the off-state. D_4, however continues to conduct and the load current is maintained. The supply voltage E_{dc} gets divided equally across the nonconducting transistors Q_1 and Q_2 hence voltage across them during mode III is ($E_{dc}/2$) volts as shown in Fig. 16.15(b).

The two-transistor forward converter is used in applications such as EPABX systems.

Advantages:

(i) Lower transistor voltage rating compared to single ended circuit.
(ii) High efficiency and reliability.
(iii) Low output ripple and noise.
(iv) Larger input transient capability.

Disadvantages:

(i) Requires two isolated driver circuits.
(ii) Poor transformer utilization compared to half bridge and push-pull configurations.
(iii) Higher cost compared to flyback circuit.
(iv) Regulation problems at light loads for multiple outputs.