

Module 3

DC to DC Converters

Lesson 23

Forward Type Switched Mode Power Supply

After completion of this lesson the reader will be able to:

- (i) Identify the topology of a forward type switched mode power supply circuit.
- (ii) Explain the principle of operation of a forward dc-to-dc power supply.
- (iii) Calculate the ratings of devices, components, transformer turns ratio for the given input and output voltages and the required output power.
- (iv) Design a simple forward type switched mode power supply circuit.

23.1 Introduction

Forward converter is another popular switched mode power supply (SMPS) circuit that is used for producing isolated and controlled dc voltage from the unregulated dc input supply. As in the case of fly-back converter (lesson-22) the input dc supply is often derived after rectifying (and little filtering) of the utility ac voltage. The forward converter, when compared with the fly-back circuit, is generally more energy efficient and is used for applications requiring little higher power output (in the range of 100 watts to 200 watts). However the circuit topology, especially the output filtering circuit is not as simple as in the fly-back converter.

Fig. 23.1 shows the **basic topology** of the forward converter. It consists of a fast switching device 'S' along with its control circuitry, a transformer with its primary winding connected in series with switch 'S' to the input supply and a rectification and filtering circuit for the transformer secondary winding. The load is connected across the rectified output of the transformer-secondary.

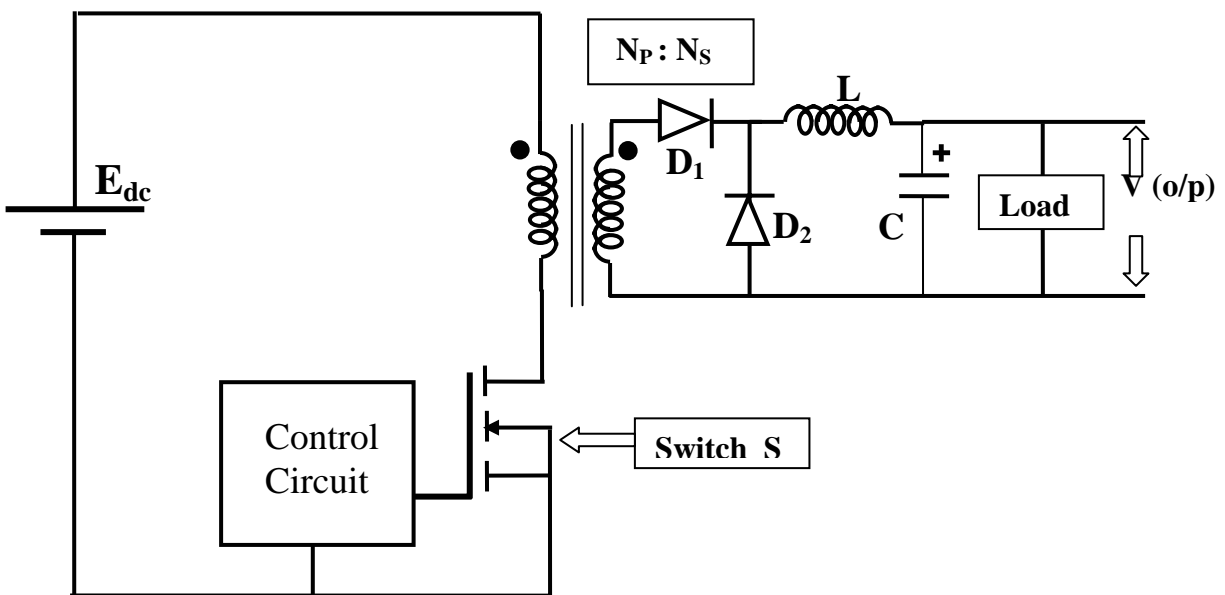


Fig. 23.1: Basic Topology of a Forward Converter

The transformer used in the forward converter is **desired to be** an ideal transformer with no leakage fluxes, zero magnetizing current and no losses. The basic operation of the circuit is explained here assuming ideal circuit elements and later the non-ideal characteristics of the devices are taken care of by suitable modification in the circuit design. **In fact, due to the presence of finite magnetizing current in a practical transformer, a tertiary winding needs to be introduced in the transformer and the circuit topology changes slightly. A**

more practical type forward converter circuit is discussed in later sections.

23.2 Principle of Operation

The circuit of Fig. 23.1 is basically a dc-to-dc buck converter with the addition of a transformer for output voltage isolation and scaling. When switch 'S' is turned on, input dc gets applied to the primary winding and simultaneously a scaled voltage appears across the transformer secondary. Dotted sides of both the windings are now having positive polarity. Diode 'D₁', connected in series with the secondary winding gets forward biased and the scaled input voltage is applied to the low pass filter circuit preceding the load. The primary winding current enters through its dotted end while the secondary current comes out of the dotted side and their magnitudes are inversely proportional to their turns-ratio. Thus, as per the assumption of an ideal transformer, the net magnetizing ampere-turns of the transformer is zero and there is no energy stored in the transformer core. When switch 'S' is turned off, the primary as well as the secondary winding currents are suddenly brought down to zero. Current through the filter inductor and the load continues without any abrupt change. Diode 'D₂' provides the freewheeling path for this current. The required emf to maintain continuity in filter-inductor current and to maintain the forward bias voltage across D₂ comes from the filter inductor 'L' itself. During freewheeling the filter inductor current will be decaying as it flows against the output voltage (V_{op}), but the presence of relatively large filter capacitor 'C' still maintains the output voltage nearly constant. The ripple in the output voltage must be within the acceptable limits. The supply switching frequency is generally kept sufficiently high such that the next turn-on of the switch takes place before the filter inductor current decays significantly. Needless to say, that the magnitudes of filter inductor and capacitor are to be chosen appropriately.

The idea behind keeping filter inductor current nearly constant is to relieve the output capacitor from supplying large ripple current. **[As per the circuit topology of Fig.23.1, the inductor and the capacitor together share the load-current drawn from the output. Under steady state condition, mean dc current supplied by the capacitor is zero but capacitor still supplies ripple current. For maintaining constant load current, the inductor and capacitor current-ripples must be equal in magnitude but opposite in sense. Capacitors with higher ripple current rating are required to have much less equivalent series resistor (ESR) and equivalent series inductor (ESL) and as such they are bulkier and costlier. Also, the ESR and ESL of a practical capacitor causes ripple in its dc output voltage due to flow of ripple current through these series impedances. Since the output voltage is drawn from capacitor terminal the ripple in output voltage will be less if the capacitor is made to carry less ripple current.]**

For better understanding of the steady-state behavior of the converter, the circuit's operation is divided in two different modes, mode-1 and mode-2. Mode-1 corresponds to the 'on' duration of the switch and mode-2 corresponds to its 'off' duration.

The following simplifying assumptions are made before proceeding to the detailed mode-wise analysis of the circuit:

- ON state voltage drops of switches and diodes are neglected. Similarly, leakage currents through the off state devices is assumed zero. The switching-on and switching-off times of the switch and diodes are neglected.

- The transformer used in the circuit is assumed to be ideal requiring no magnetizing current, having no leakage inductance and no losses.
- The filter circuit elements like, inductors and capacitors are assumed loss-less.
- For the simplified steady-state analysis of the circuit the switch duty ratio (δ), as defined in the previous chapters is assumed constant.
- The input and output dc voltages are assumed to be constant and ripple-free. Current through the filter inductor (L) is assumed to be continuous.

Mode-1 of Circuit Operation

Mode-1 of circuit starts after switch 'S' (as shown in Fig.23.1) is turned ON. This connects the input voltage, E_{dc} , to the primary winding. Both primary and secondary windings start conducting simultaneously with the turning on of the switch. The primary and secondary winding currents and voltages are related to their turns-ratio (N_P / N_S), as in an ideal transformer.

Fig.23.2 (a) shows, in bold lines, the current carrying path of the circuit and Fig.23.2 (b) shows the functional equivalent circuit of mode-1. As switch 'S' closes, diode D_1 in the secondary circuit gets forward biased and the input voltage, scaled by the transformer turns ratio, gets applied to the secondary circuit. Diode D_2 does not conduct during mode-1, as it remains reverse biased.

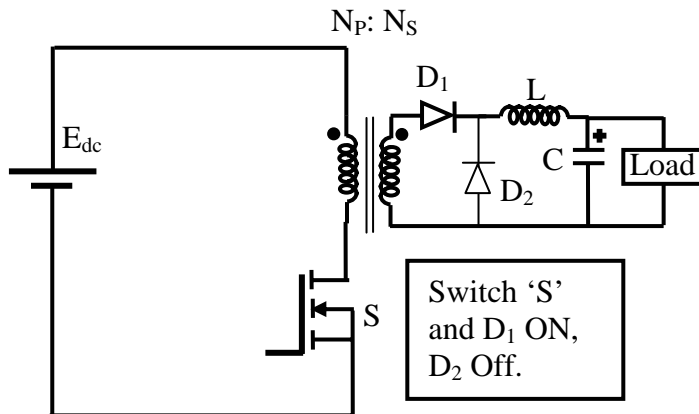


Fig. 23.2(a): Current path during Mode-1

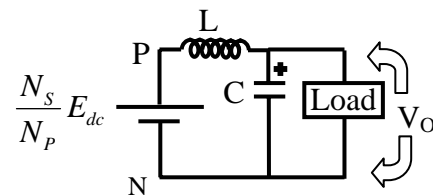


Fig.23.2(b): Equivalent circuit in Mode-1

As can be seen, the output circuit consisting of L-C filter and the load gets a voltage equal to $\frac{N_S}{N_P} E_{dc}$ during mode-1. This voltage is shown across points 'P' and 'N' in Fig. 23.2(b) and it is

the maximum achievable dc voltage across the load, corresponding to $\delta = 1$. Mode-1 can be called as powering mode during which input power is transferred to the load. Mode-2, to be called as freewheeling mode, starts with turning off of the switch 'S'.

Mode-2 of Circuit Operation

As soon as switch 'S' is turned off, the primary and the secondary winding currents of the transformer fall to zero. However, the secondary side filter inductor maintains a continuous current through the freewheeling diode ' D_2 '. Diode ' D_1 ' remains off during this mode and isolates the output section of the circuit from the transformer and the input.

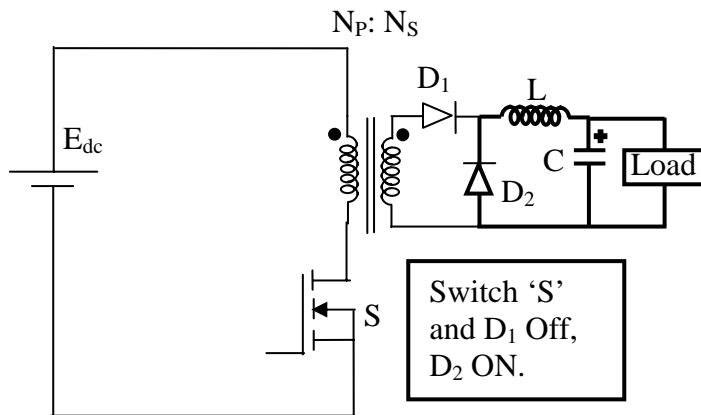


Fig. 23.3(a): Current path during Mode-2

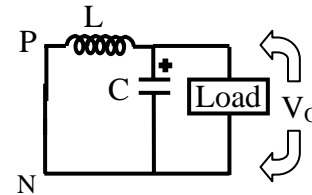


Fig.23.2(b): Equivalent circuit in Mode-2

Fig. 23.3(a) shows the current carrying portion of the circuit in bold line and Fig. 23.3(b) shows the equivalent circuit active during mode-2. Points 'P' and 'N' of the equivalent circuit are effectively shorted due to conduction of diode 'D₂'. The inductor current continues to flow through the parallel combination of the load and the output capacitor. During mode-2, there is no power flow from source to load but still the load voltage is maintained nearly constant by the large output capacitor 'C'. The charged capacitor and the inductor provide continuity in load voltage. However since there is no input power during mode-2, the stored energy of the filter inductor and capacitor will be slowly dissipating in the load and hence during this mode the magnitudes of inductor current and the capacitor voltage will be falling slightly. In order to keep the load voltage magnitude within required tolerance band, the converter-switch 'S' is turned on again to end the freewheeling mode and start the next powering mode (mode-1). Under steady state, loss in inductor current and capacitor voltage in mode-2 is exactly made up in mode-1. It may not be difficult to see that to maintain load voltage within the desired tolerance band the filter inductor and capacitor magnitudes should be sufficiently large. However, in order to keep the filter cost and its physical size small these elements should not be unnecessarily too large. Also, for faster dynamic control over the output voltage the filter elements should not be too large. **[It may be pointed out here that the filter inductor, capacitor, transformer and the heat sinks for the switching devices together account for nearly 90% of the power supply weight and volume.]**

One important factor that directly influences the size of the filter circuit elements and the transformer is the converter's switching frequency. High frequency operation of switch 'S' will help in keeping the filter and transformer size small. The switching frequency of a typical forward converter may thus be in the range of 100 kHz or more. The higher end limit on the switching frequency comes mainly due to the finite switching time and finite switching losses of a practical switch.

Switch limitations have been ignored in the simplified analysis presented here. As mentioned earlier, the switch and the diodes have been assumed to be ideal, with no losses and zero switching time. **Control over switch duty ratio, which is the ratio of ON time to (ON + OFF) time, provides the control over the output voltage 'V_o'.**

Relation Between Input and Output Voltage

The equivalent circuits of mode-1 and mode-2 can be used to derive a steady state relation between the input voltage, switch duty ratio (δ) and the output voltage. With the assumption of

constant input and output voltage, the instantaneous value of inductor voltage (e_L) during mode-1 can be written as:

$$e_L(t) = \frac{N_s}{N_p} E_{dc} - V_O; \text{ for } 0 \leq t \leq \delta T, \quad \text{----- (23.1)}$$

Where $t = 0$ is the time instant when mode-1 of any steady state switching cycle starts, T is the switching time period that may be assumed to be constant and δ is the duty ratio of the switch. It can be seen that δT is the time duration of mode-1 and $(1-\delta) T$ is the time duration of mode-2. The inductor voltage during mode-2 may similarly be written as:

$$e_L(t) = -V_O; \text{ for } \delta T \leq t \leq T, \quad \text{----- (23.2)}$$

Now since voltage across an inductor, averaged over a steady state cycle time, must always be zero, one gets:

$$\left[\frac{N_s}{N_p} E_{dc} - V_O \right] \delta + [-V_O] (1-\delta) = 0,$$

$$\text{Or, } V_O = \delta \frac{N_s}{N_p} E_{dc} \quad \text{----- (23.3)}$$

Thus according to Eqn. (23.3), the forward converter output voltage is directly proportional to the switch duty ratio. It may be noticed that except for transformer scaling factor the output voltage relation is same as in a simple dc-to-dc buck converter. **It is to be noted that the output voltage relation given by Eqn. (23.3) is valid only under the assumption of continuous inductor current.** For an improperly designed circuit or for very light load at the converter output, the inductor current may decay to zero in the midst of mode-2 resulting into discontinuous inductor current. Once the inductor current becomes zero, diode 'D₂' in Fig. 23.3(a) no longer conducts and the points 'P' and 'N' of the equivalent circuit in Fig. 23.3(b) are no longer shorted. In fact, the output voltage 'V_O' will appear across 'P' and 'N'. Thus equation (23.2) remains valid only for a part of $(1-\delta) T$ period. In case of discontinuous inductor current, the output voltage, which is the average of voltage across points 'P' and 'N' will have a higher magnitude than the one given by Eqn. (23.3). **Under discontinuous inductor current the relation between output voltage and switch duty ratio becomes non-linear and is load dependent.** For better control over output voltage discontinuous inductor current mode is generally avoided. With prior knowledge of the load-range and for the desired switching frequency the filter inductor may be suitably chosen to keep the inductor current continuous and preferably with less ripple.

23.3 Practical Topology of A Forward Converter Circuit

Fig. 23.4 shows the circuit topology of a practical forward converter. It takes into account the non-ideal nature of a practical transformer. Other non-idealities of the circuit elements like that of switch, diodes, inductor and capacitor are taken care of by modifying the circuit parameters chosen on the basis of ideal circuit assumption. Most common consequence of non-ideal nature of circuit elements is increase in losses and hence reduction in efficiency of the power supply. A practical way to get around the consequence of circuit losses is to over-design the power supply.

The design should aim to achieve an output power of $\frac{P_o}{\eta}$, where 'P_o' is the required output power and 'η' is the efficiency of the converter. As a first order approximation, a typical efficiency figure of around 80% may be assumed for the forward converter. Once the efficiency figure has been considered the circuit may still be designed based on the simplified analysis presented here,

which neglects many of the non-idealities. Another common non-ideality is the low frequency ripple and fluctuation in input dc supply voltage. In the simplified analysis input supply has been assumed to be of constant magnitude. In a practical circuit, the variation in input supply is taken care of by modulating the switch duty ratio in such a manner that it offsets the effect of supply voltage fluctuation and continues to give the required quality of output voltage.

The non-ideality of the transformer, however, cannot simply be overcome by changing the circuit parameters of the simplified circuit shown in Fig. 23.1. A practical transformer will have finite magnetization current and finite energy associated with this magnetization current. Similarly there will be some leakage inductance of the windings. However, windings of the forward-converter transformer will have much smaller leakage inductances than those of fly-back converter transformer. In fly-back transformer's flux path some air-gap is deliberately introduced by creating a gap in the transformer core (refer to lesson-22). Introduction of air gap in the mutual flux path increases the magnitude of leakage inductances. Transformer of a forward converter should have no air-gap in its flux path.

The forward-converter transformer works like a normal power transformer where both primary and secondary windings conduct simultaneously with opposing magnetomotive force (mmf) along the mutual flux path. The difference of the mmfs is responsible for maintaining the magnetizing flux in the core. When primary winding current is interrupted by switching off 'S', the dotted ends of the windings develop negative potential to oppose the interruption of current (in accordance with Lenz's law). Negative potential of the dotted end of secondary winding makes diode 'D₁' reverse biased and hence it also stops conducting. This results in simultaneous opening of both primary and secondary windings of the transformer. In case the basic circuit of Fig. 23.1 is used along with a practical transformer, turning off of switch 'S' will result in sudden demagnetization of the core from its previously magnetized state. As discussed in Lesson-22, a practical circuit cannot support sudden change in flux. Any attempt to change flux suddenly results in generation of infinitely large magnitude of voltage (in accordance with Lenz's law). Such a large voltage in the circuit will have a destructive effect and that should be avoided. Thus, after switch 'S' is turned off, there must exist a convenient path for the trapped energy in the primary due to magnetizing current. One solution could be a snubber circuit across the primary winding, similar to the one shown in Fig.22.6 for a fly-back circuit (refer to lesson-22). Each time the switch 'S' is turned off the snubber circuit will dissipate the energy associated with the magnetizing flux. This, as has been seen in connection with fly-back converter, reduces the power-supply efficiency considerably. A more preferred solution is to recover this energy. For this reason the practical forward converter uses an extra tertiary winding with a series diode, as shown in Fig. 23.4. When both switch 'S' and 'D₁' turn-off together, as discussed above, the magnetization energy will cause a current flow through the closely coupled tertiary winding and the diode 'D₃'. The dot markings on the windings are to be observed. Current entering the dot through any of the magnetically coupled windings will produce magnetic flux in the same sense. As soon as switch 'S' is turned off, the dotted end voltages of the windings will become negative in accordance with Lenz's law. The sudden rise in magnitude of negative potential across the windings is checked only by the conduction of current through the tertiary winding. As discussed earlier unless the continuity in transformer flux is maintained the voltages in the windings will theoretically reach infinite value. Thus turning off of switch 'S' and turn-on of diode 'D₃' need to be simultaneous. Similarly fall in magnetizing current through primary winding must be coupled with simultaneous rise of magnetization current through the tertiary winding. In order that the entire flux linking the primary winding gets transferred to the tertiary, the magnetic coupling between these two windings must be very good. For this the primary and tertiary winding turns

are wound together, known as bifilar windings. The wires used for bifilar windings of the primary and the tertiary need to withstand large electrical voltage stress and are costlier than ordinary transformer wires.

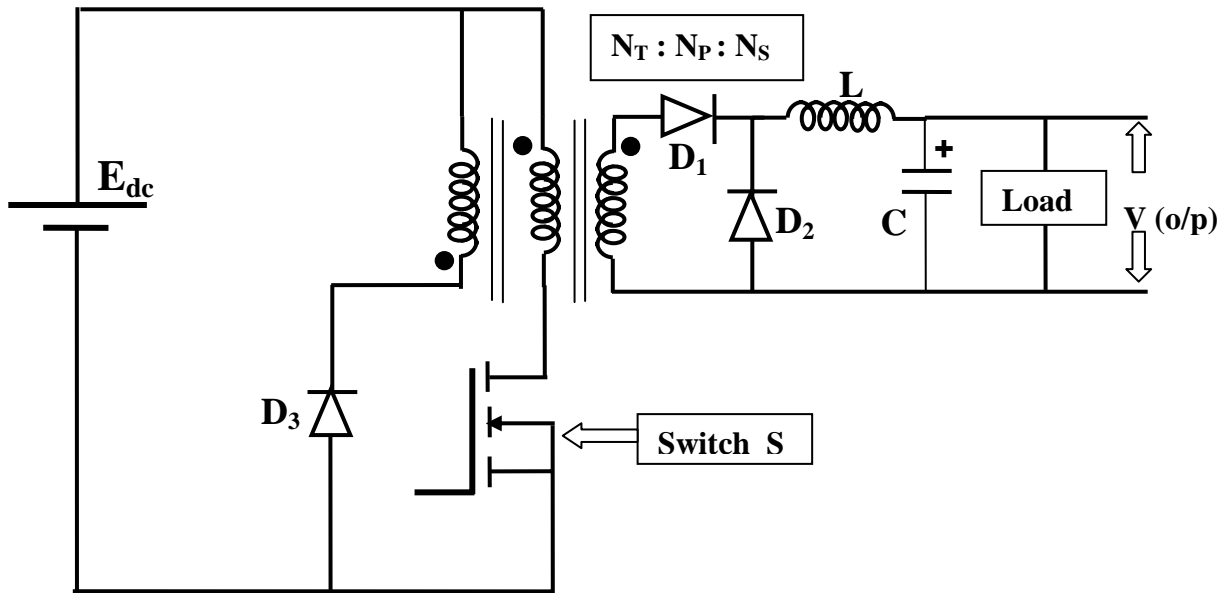


Fig. 23.4: Circuit topology of a practical forward converter

Fig. 23.5 shows some of the typical current and voltage waveforms of the forward converter shown in Fig. 23.4. For these waveforms, once again, many of the ideal circuit assumptions have been made.

In Fig. 23.5, V_{load} is the converter output voltage that is maintained constant at V_O . ' I_L ' is the current through filter inductor ' L '. The inductor current rises linearly during mode-1 as its voltage is maintained constant as per Eqn.23.1. Similarly the inductor current decays at a constant rate in mode-2 as it flows against the constant output voltage. Average magnitude of inductor current equals the load current. I_{SW} and V_{SW} are respectively the switch current and switch voltage. V_{D3} is the voltage across diode ' D_3 '. Switch conducts only during mode-1 and carries the primary winding current (I_{Pr}) of the transformer. The transformer magnetization current is assumed to be negligibly small and hence the primary winding essentially carries the reflected inductor current.

As switch ' S ' turns on, primary winding gets input dc voltage (with its dotted end positive). The induced voltages in other windings are in proportion to their turns ratios. Diode ' D_3 ' of the tertiary winding is reverse biased and is subjected to a voltage $-E_{dc} \left(1 + \frac{N_T}{N_P}\right)$.

As soon as switch ' S ' is turned-off, primary and secondary winding currents fall to zero but diode ' D_3 ' gets forward biased and the tertiary winding starts conducting to maintain a path for the magnetizing current. While ' D_3 ' conducts the tertiary winding voltage is clamped to input dc voltage with its dotted end negative. Primary and secondary windings have induced voltages due to transformer action. Primary winding voltage equals to $E_{dc} \frac{N_P}{N_T}$, with dotted end at negative potential.

In Fig. 23.5, ' V_{Pr} ' denotes the primary winding voltage. The net volt-time area of the primary

winding voltage must be zero under steady state. Voltage across switch 'S' can be seen to be the sum of primary winding voltage and the input voltage and equals $E_{dc}(1 + \frac{N_p}{N_T})$.

As the tertiary winding current flows against the input dc supply, the magnetization current decays linearly given by the following relation:

$$N_T \frac{d\Phi_m}{dt} = -E_{dc} \quad \text{----- (23.4)}$$

Where, Φ_m = flux through the transformer core.

When the transformer is completely demagnetized, diode 'D₃' turns off and voltage across transformer windings fall to zero. The transformer remains de-magnetized for the remaining duration of Mode-2.

When switch 'S' is again turned on, in the next switching cycle, the transformer flux builds up

linearly given by the relation: $N_p \frac{d\Phi_m}{dt} = E_{dc}$, ----- (23.5)

Under steady-state the increase in flux during conduction of switch 'S' must be equal to fall in flux during conduction of tertiary winding and hence Eqns. 23.4 and 23.5 may be combined to

show that $\frac{t_T}{N_T} = \frac{t_p}{N_p}$, where t_T and t_p are the time durations for which tertiary and primary

windings conduct during each switching cycle. Now, $t_p = \delta T$ = on-duration of switch 'S' and the tertiary winding conducts only during off duration of switch (during mode-2). Hence, $(1-\delta) T \geq t_T$. As a result,

$$\frac{\delta}{1-\delta} \leq \frac{N_p}{N_T} \quad , \text{ or } \delta \leq \frac{N_p}{(N_p + N_T)} \quad \text{----- (23.6)}$$

Thus if $N_p = N_T$, the duty ratio must be less than or equal to 50% or else the transformer magnetic circuit will not get time to reset fully during mode-2 and will saturate. Less duty ratio means less duration of powering mode (mode-1) and hence less transfer of power to the output circuit. On the other hand, as described above, if $\frac{N_p}{N_T}$ is increased for higher duty ratio, the switch voltage stress increases.

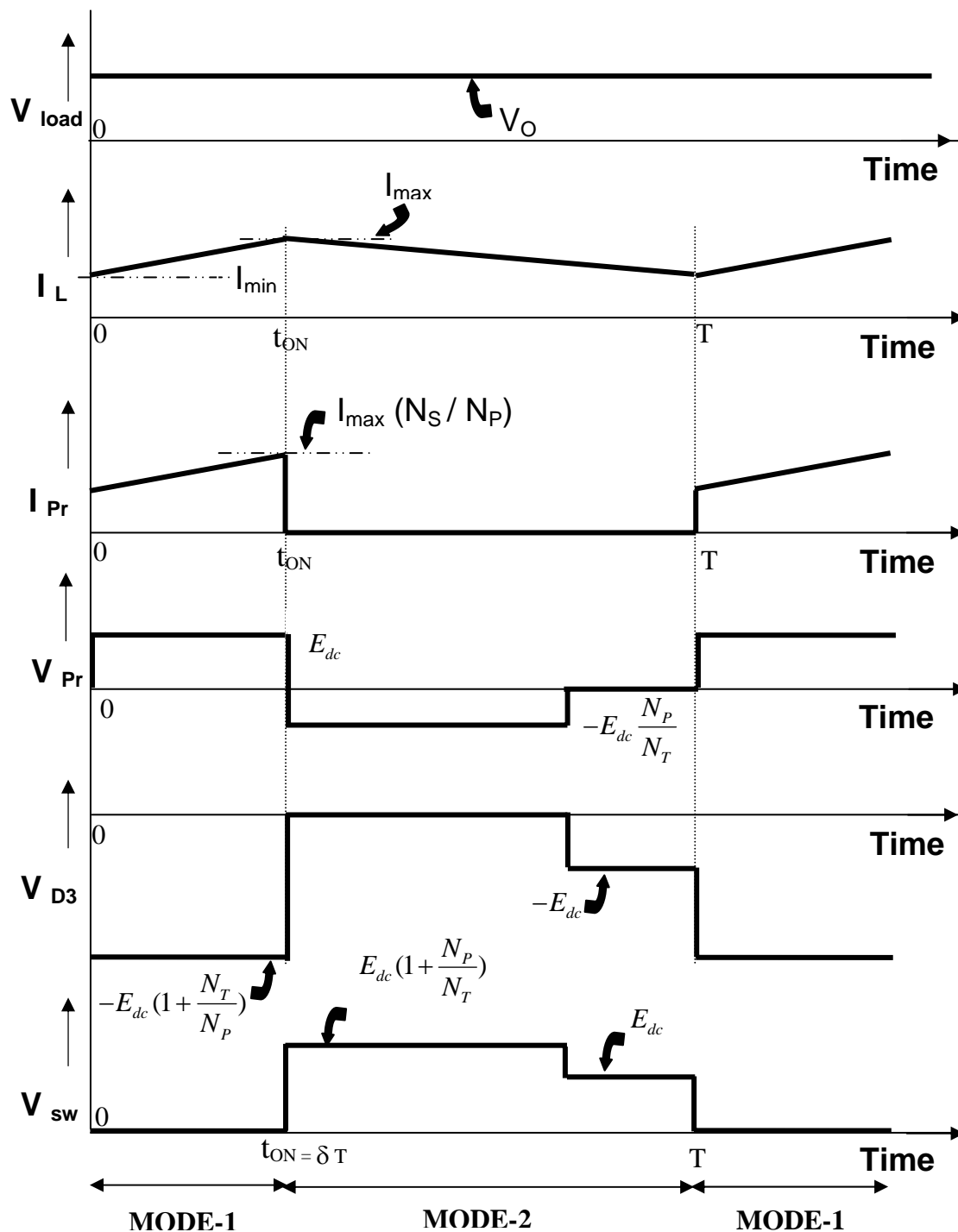


Fig.23.5: Some Typical waveforms of a practical Forward converter circuit

23.4 Selection of Transformer Turns Ratio

The transformer-winding turns ratio is a crucial design factor. The primary to secondary turns ratio of the transformer is decided in accordance with Eqn. 23.3. For the required output voltage (V_O), the turns ratio $\frac{N_S}{N_P}$ is found after considering the minimum magnitude of input supply

voltage (E_{dc}) and the maximum allowable duty ratio (δ). The maximum duty ratio of the converter, as discussed above, is constrained by the primary to tertiary winding turns ratio (given by Eqn. 23.6) but the choice of primary to tertiary winding turns ratio is often governed by the voltage stress that the switch must withstand. Higher voltage stress will mean higher cost of switch. If the tertiary winding turns is kept very high, the switch voltage stress reduces but allowable duty ratio of switch and the power output of the converter becomes low and diode 'D₃' voltage rating increases. Thus an optimum design needs to be arrived at to maximize the performance of the converter.

23.5 Selection of Filter Circuit Inductor and Capacitor

The transformer's secondary voltage is rectified and filtered suitably to get the desired quality of output voltage waveform. The filter inductor and capacitor values need to be chosen optimally to arrive at a cost-effective, less bulky power supply. In this section, some simple guidelines have been developed to arrive at the required filter size.

Inductor current waveform during a typical switching cycle has been shown in Fig. 23.5. As described earlier, mean (dc) value of inductor current equals the load current. The filter capacitor merely supplies the ripple (ac) current of switching frequency. It has also been mentioned earlier that for linear relation between the output voltage and the switch duty ratio, the inductor current is desired to be continuous (refer to Eqn.23.3). In case the inductor current becomes discontinuous the linearity between switch duty ratio and output voltage is lost and the output-voltage controller circuit, which is often designed using linear control theory, is not able to maintain the desired quality of output voltage. Hence filter inductor should be chosen to be sufficiently large such that under expected range of load current variation, the inductor current remains continuous. In many cases the minimum value of load current may not be specified or may be too low. If the load connected to the output is very light or if there is no load, the inductor current will not remain continuous. Hence, as a thumb rule, the filter inductor size may be chosen such that the inductor current remains continuous for more than 10% of the rated load current. At 10% of the load, the inductor current may be assumed to be just continuous. This gives a basis for choosing the inductor value as detailed below:

With reference to the waveforms in Fig. 23.5, under just continuous inductor current, $I_{\min} = 0$ and

$$I_{\text{load}} = 0.5 (I_{\min} + I_{\max}) = 0.5 I_{\max} = 0.1 I_{\text{rated}}, \text{ where } I_{\text{rated}} \text{ is the rated load current.}$$

$$\text{Again, using Eqn.23.1, } (I_{\max} - I_{\min}) = \delta T \left(\frac{N_s}{N_p} E_{dc} - V_O \right) / L \text{ ----- (23.7)}$$

Thus for $I_{\min} = 0$ and $I_{\max} = 0.2 I_{\text{rated}}$,

$$L = \delta T \left(\frac{N_s}{N_p} E_{dc} - V_O \right) / (0.2 I_{\text{rated}}) \text{ ----- (23.8)}$$

where V_O , the output voltage, is assumed to have a fixed magnitude. Input supply voltage, E_{dc} , may itself be varying and the duty ratio is adjusted to keep V_O constant in accordance with Eqn.23.3. Thus even though ' E_{dc} ' and ' δ ' are varying, their product $\left(\delta \frac{N_s}{N_p} E_{dc} \right)$ will be constant

and equal to V_O . [As mentioned earlier, only low frequency variation in supply voltage has been considered. Switching frequency and the switch control dynamics are assumed to be much faster.]

Hence, the inductor 'L' magnitude should correspond to minimum value of duty ratio and may be

$$\text{written as } L = \frac{5V_O}{I_{rated} f_{SW}} (1 - \delta_{min}) \quad \text{----- (23.9)}$$

, where δ_{min} is the minimum magnitude of duty ratio and f_{SW} is the constant switching frequency of the converter switch. Now in accordance with Eqn.(23.6) the maximum value of duty ratio may be taken as $\delta_{max} = \frac{N_P}{(N_P + N_T)}$. Again to maintain constant output voltage

$$\frac{\delta_{max}}{\delta_{min}} = \frac{E_{dc,max}}{E_{dc,min}} \quad \text{----- (23.10)}$$

, where $E_{dc,max}$ and $E_{dc,min}$ are maximum and minimum magnitudes of input dc voltage respectively.

Thus $\delta_{min} = \frac{E_{dc,min}}{E_{dc,max}} \frac{N_P}{(N_P + N_T)}$ and

$$L = \frac{5V_O}{I_{rated} f_{SW}} \left[1 - \frac{E_{dc,min}}{E_{dc,max}} \frac{N_P}{(N_P + N_T)} \right] \quad \text{----- (23.11)}$$

The inductor magnitude given by Eqn.(23.11) will limit the worst case peak to peak current ripple in the filter inductor ($= I_{max} - I_{min}$) to 20% of rated current. [refer to Eqns.(23.7) and (23.8). It may be noted here that as long as inductor current is continuous the peak-to-peak ripple in the inductor current is not affected by the dc value of load current. For constant output voltage and constant current through load, the inductor current ripple depends only on the duty ratio, which in turn depends on the magnitude of input dc voltage]

Once inductor magnitude is chosen in accordance with Eqn.(23.11), peak to peak ripple in the capacitor current will also be 20% of the rated current. This is so because the load, under steady state, has been assumed to draw a constant magnitude of current.

Even though the output capacitor voltage has been assumed constant in our analysis so far, there will be a minor ripple in capacitor voltage too which however will have only negligible effect on the analysis carried out earlier. The worst case, peak to peak ripple in capacitor voltage

$$(v_{O,p-p}) \text{ can be given as: } v_{O,p-p} = \frac{I_{rated}}{20Cf_{SW}} \quad \text{----- (23.12)}$$

, where 'C' is the output capacitance in farad. Capacitance value should be chosen, in accordance with the above equation, based on the allowed ripple in the output voltage.

Quiz

1). If the turns ratio of the primary and tertiary windings of the forward transformer are in the ratio of 1:2, what is the maximum duty ratio at which the converter can be operated? Corresponding to this duty ratio, what should be the minimum ratio of secondary to primary

turns if the input dc supply is 400 volts and the required output voltage is 15 volts? Neglect switch and diode conduction voltage drops.

[Answer: 1/3 and 9/80]

2) Find maximum voltage stress of the switch in the primary winding and diode in the tertiary winding if the converter-transformer has 10 primary turns and 15 tertiary turns and the maximum input dc voltage is 300 volts.

[Answer: Switch voltage stress = 500V, diode voltage stress = 750V]

3) Calculate the filter inductor and capacitor values for the forward converter described below: Maximum duty ratio = 0.5, Input dc remains constant at 200 volts, output dc (under steady state) = 10 volts \pm 0.1 volt, primary to secondary turns = 10:1. The load current is expected to vary between 0.5 and 5 amps. Assume just continuous conduction of inductor current at 0.5 amp load current. Take switching frequency = 100 kHz.

[Answer: L = 50 micro Henry and C = 12.5 micro Farad]

(4) What function does the diode 'D₁' of circuit in Fig.(23.4) have?

- (i) rectifies secondary voltage
- (ii) blocks back propagation of secondary voltage to transformer
- (iii) both (i) and (ii)
- (iv) protects diode 'D₂' from excessive reverse voltage

[Answer: (iii)]