DC-DC Converters

http://www.powere.dynamictopway.com/index.htm

Modern electronic systems require high-quality, small, lightweight, reliable, and efficient power supplies.

High-frequency electronic power processors are used in dc-dc power conversion. The functions of dc-dc converters are:

1- to convert a dc input voltage V_S into a dc output voltage V_O;
2- to regulate the dc output voltage against load and line variations;
3- to reduce the ac voltage ripple on the dc output voltage below the required level;
4- to provide isolation between the input source and the load (isolation is not always required);
5- to protect the supplied system and the input source from electromagnetic interference (EMI); and
6- to satisfy various international and national safety standards.

The dc-dc converters can be divided into two main types:

a- **hard-switching pulsewidth modulated (PWM) converters,**
b- **and resonant and soft-switching converters.**

This chapter deals with PWM dc-dc converters, which have been very popular for the last three decades, and that are widely used at all power levels. Topologies and properties of PWM converters are well understood and described in the literature. Advantages of PWM converters include low component count, high efficiency, constant frequency operation, relatively simple control and commercial availability of integrated circuit controllers, and ability to achieve high conversion ratios for both stepdown and step-up application. A disadvantage of PWM dc-dc converters is that PWM rectangular voltage and current waveforms cause turn-on and turn-off losses in semiconductor devices, which limit practical operating frequencies to hundreds of kilohertz. Rectangular waveforms also inherently generate EMI.

DC Choppers:

A step-down dc chopper with a resistive load is shown in Fig.13.1a. It is a series connection of a dc input voltage source VS, controllable switch S, and load resistance R. In most cases, switch S has unidirectional voltage-blocking capabilities and unidirectional current-conduction capabilities. Power electronic switches are usually implemented with power MOSFETs, IGBTs, MCTs, power BJTs, or GTOs.

The switch is being operated with a duty ratio D defined as a ratio of the switch on time to the sum of the on and off times.

For a constant frequency operation

$$D = \frac{t_{on}}{t_{on} + t_{off}} = \frac{t_{on}}{T} \qquad (13.1)$$

where T =1/f is the period of the switching frequency f .

The average value of the output voltage is

$$V_O = DV_S \qquad (13.2)$$

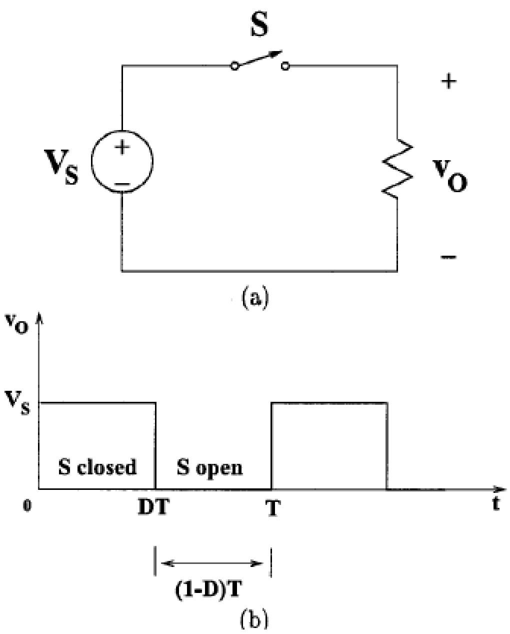


Figure 13.1 DC chopper with resistive load:
(a) circuit diagram; (b) output voltage waveform.

The dc step-down choppers are commonly used in dc drives. In such a case, the load is represented as a series combination of inductance L, resistance R, and back-emf E as shown in Fig. 13.2a. To provide a path for a continuous inductor current flow when the switch is in the off state, an antiparallel diode D must be connected across the load.

Because the chopper of Fig. 13.2a provides a positive voltage and a positive current to the load, it is called a first-quadrant chopper. The load voltage and current are graphed in Fig.13.2b under assumptions that the load current never reaches zero and the load time constant

$\tau = L/R$

is much greater than the period T. Average values of the output voltage and current can be adjusted by changing the duty ratio *D*.

The dc choppers can also provide peak output voltages higher than the input voltage. Such a step-up configuration is presented in Fig. 13.3. It consists of dc input source V_S, inductor L connected in series with the source, switch S connecting the inductor to ground, and a series combination of diode D and load. If the switch operates with a duty ratio *D*, the output voltage is a series of pulses of duration

$(1 - D)T$ and amplitude $V_S / (1 - D)$.

Therefore, neglecting losses, the average value of the output voltage is V_S. To obtain an average value of the output voltage greater than V_S, a capacitor must be connected in parallel with the load.

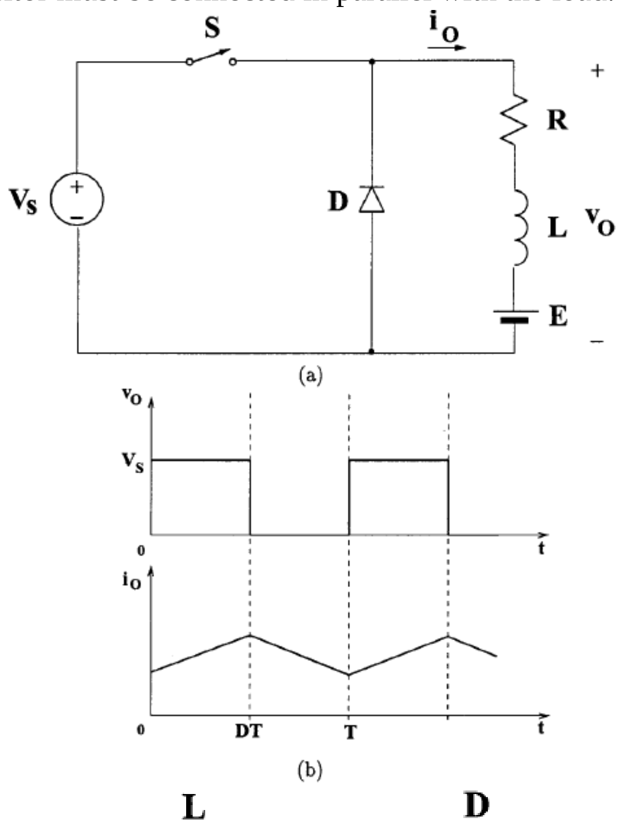


Figure 13.2 DC chopper with RLE load:
(a) circuit diagram; (b) waveforms.

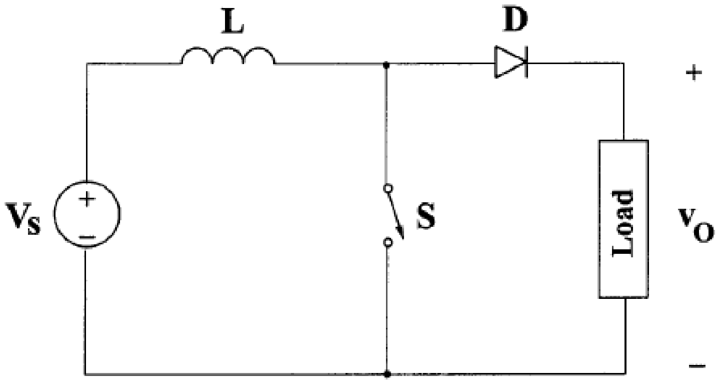


Figure 13.3 The dc step-up chopper.

Step Down DC/DC Converter Chopper (Buck Converter):

Basic Converter

The step-down dc-dc converter, commonly known as a buck converter, is shown in Fig. 13.4a. It consists of dc input voltage source V_S, controlled switch S (a thyristor and its commutation circuit or a power transistor IGBT or MOSFET), diode D, filter inductor L, filter capacitor C, and load resistance R.

The following abbreviations are used in the analysis:

(CCM): Continuous Conduction Mode
(DCM): Discontinuous Conduction Mode
$\tau = L/R$: Load time constant
$T = 1/f$: Duty cycle time or modulation cycle time
$D = T_{ON}/T$: Time ratio
V_r = peak to peak ripple voltage
Im = E/R: Maximum ON-state load current (i.e., when $D = 1$).

The relationship among the input voltage, output voltage, and the switch duty ratio *D* can be derived, for instance, from the inductor voltage vL waveform (see Fig. 13.4b). According to **Faraday's law**, the inductor volt-second product over a period of steady-state operation is zero. For the buck converter

(Vs-Vo) ton = - Vo toff

$$(V_S - V_O)DT = -V_O(1 - D)T \qquad (13.3)$$

Hence, the dc voltage **transfer function** M_V

$$M_V = \frac{V_O}{V_S} = D \qquad (13.4)$$

The dc-dc converters can operate in two distinct modes with respect to the inductor current i_L. Figure 13.4b depicts the CCM in which the inductor current is always greater than zero. When the average value of the output current is low (high R) and/or the switching frequency f is low, the converter may enter the discontinuous conduction mode (DCM) In the DCM, the inductor current is zero during a portion of the switching period.

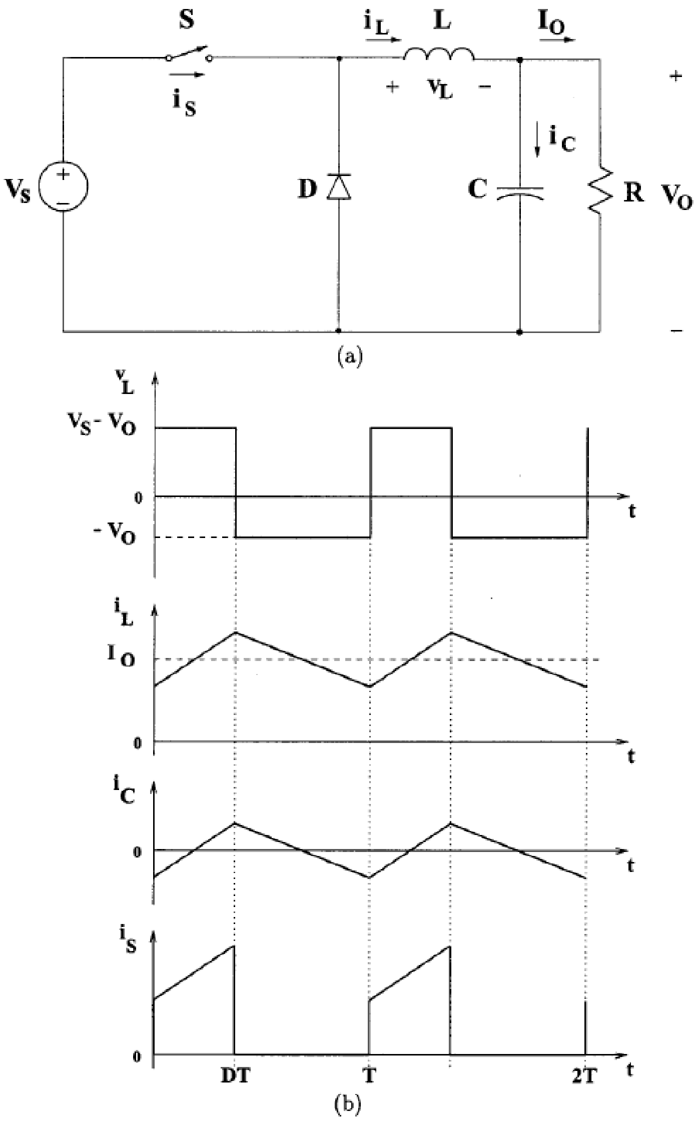


Figure 13.4 Buck converter: (a) circuit diagram; (b) waveforms.

It is uncommon to mix these two operating modes because of different control algorithms. For the buck converter, the value of the filter inductance that determines the boundary between CCM and DCM is given by

$$L_b = \frac{(1-D)R}{2f} \qquad (13.5)$$

Ex. 1:
For typical values of D = 0.5, R = 10Ω, and f = 100 kHz,

$$L_b = \frac{(1-D)R}{2f} = \text{-----------------} = 25\ \mu\text{H}$$

the boundary is *Lb* = 25 μH. For L > Lb, the converter operates in the CCM.

Almost all of this ac component flows through the filter capacitor as a current i_C . Current i_C causes a small voltage ripple across the dc output voltage V_O. To limit the peak-topeak value of the ripple voltage below a certain value *Vr* , the filter capacitance C must be greater than

$$C_{\min} = \frac{(1-D)V_O}{8V_r L f^2} \qquad (13.6)$$

Ex. 2:
At *D* = 0.5, Vr / V_O =1%, L =25 μH, and f = 100 kHz, therefore minimum capacitance is

$$C_{\min} = \frac{(1-D)V_O}{8V_r L f^2} = \text{-------------------------} = 25\ \mu\text{F}$$

$\therefore\ C_{\min} = 25\ \mu\text{F}$.

<u>Notes for the Design:</u>

1- The designer needs to determine values of passive components L and C, and of the switching frequency f .
2- The switching frequency is limited, however, by the type of semiconductor switches used and by switching losses.
3- Eqns (13.5) and (13.6) are the key design equations for the buck converter and it can be accomplished by using a high switching frequency f ..
4- The input and output dc voltages (hence, the duty ratio D)
5- the range of load resistances R are usually determined by preliminary specifications.
6- The value of the filter inductor L is calculated from the CCM/DCM condition using Eq. (13.5).
7- The value of the filter capacitor C is obtained from the voltage ripple condition Eq. (13.6).
8- For the compactness and low conduction losses of a converter, it is desirable to use small passive components.

Transformer Versions of Buck Converter

In many dc power supplies, an isolation between the dc or ac input and the dc output is required for safety and reliability. An economical means of achieving such an isolation is to employ a transformer **(Pulse Transformers)** version of a dc-dc converter. High frequency transformers are of a small size and low weight and provide high efficiency. Their turns ratio can be used additionally to adjust the output voltage level.

The most popular types of buckderived dc-dc converters are

1. Forward converter
2. Push-pull converter
3. Half-bridge converter
4. Full-bridge converter

1- Forward Converter

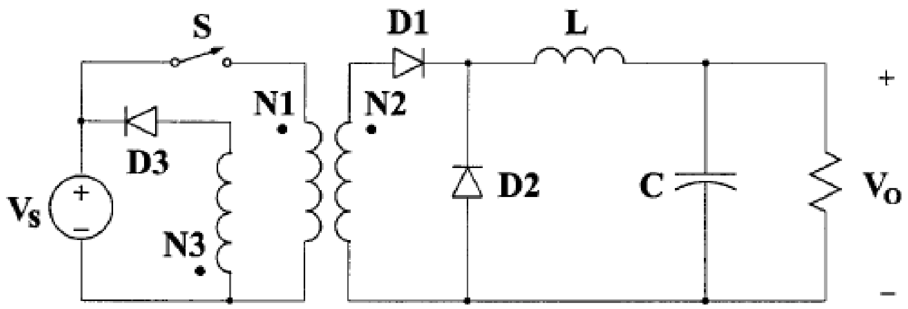


Figure 13.5 Forward converter.

The circuit diagram of a forward converter is depicted in Fig.13.5. When the switch S is on, diode D1 conducts and diode D2 is off. The energy is transferred from the input, through the transformer, to the output filter. When the switch is off, the state of diodes D1 and D2 is reversed. The dc voltage transfer function of the forward converter is

$$M_V = \frac{D}{n} \qquad (13.7)$$

where n = N1/N2.

In the forward converter, the energy-transfer current flows through the transformer in one direction. Hence, an additional winding with diode D3 is needed to bring the magnetizing current of the transformer to zero, which prevents transformer saturation. The turns ratio N1=N3 should be selected in such a way that the magnetizing current decreases to zero during a fraction of the time interval when the switch is off.

Eqns (13.5) and (13.6) can be used to design the filter components. The forward converter is very popular for lowpower applications.

For medium-power levels, converters with bidirectional transformer excitation (push-pull, half-bridge, and full-bridge) are preferred due to better utilization of magnetic components.

2- Push-Pull Converter

The PWM dc-dc push-pull converter is shown in Fig. 13.6. The switches S1 and S2 operate shifted in phase by T/2 with the same duty ratio D, however, the duty ratio must be smaller than 0.5. When switch S1 is on, diode D1 conducts and diode D2 is off; the diode states are reversed when switch S2 is on. When both controllable switches are off, the diodes are on and share equally the filter inductor current. The dc voltage transfer function of the push-pull converter is

$$M_V = \frac{2D}{n} \qquad (13.8)$$

Where n = N1/N2

The boundary value of the filter inductor is

$$L_b = \frac{(1-2D)R}{4f} \qquad (13.9)$$

The filter capacitor can be obtained from

$$C_{\min} = \frac{(1-2D)V_O}{32V_rLf^2} \qquad (13.10)$$

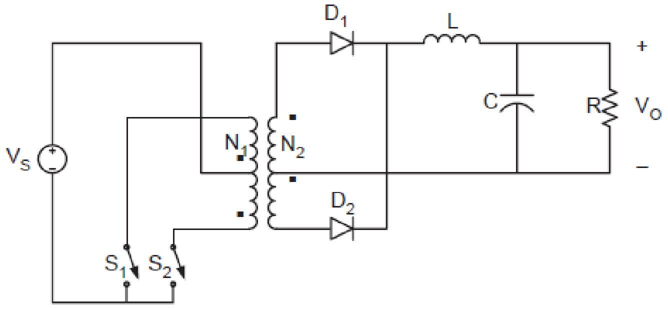


Figure 13.6 Push–pull converter.

3- Half-Bridge Converter

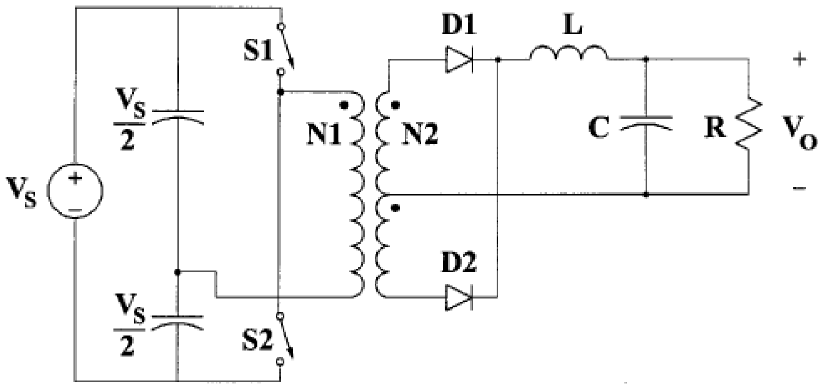


Figure 13.7 Half-bridge converter.

Figure 13.7 shows the dc-dc half-bridge converter. The operation of the PWM half-bridge converter is similar to that of the push-pull converter. In comparison to the push-pull converter, the primary switch of the transformer is simplified at the expense of two voltage-sharing input capacitors. The halfbridge converter dc voltage transfer function is

$$M_V \equiv \frac{V_O}{V_S} = \frac{D}{n} \qquad (13.11)$$

where $D \le 0.5$

Equations (13.9) and (13.10) apply to the filter components.

4- Full-Bridge Converter

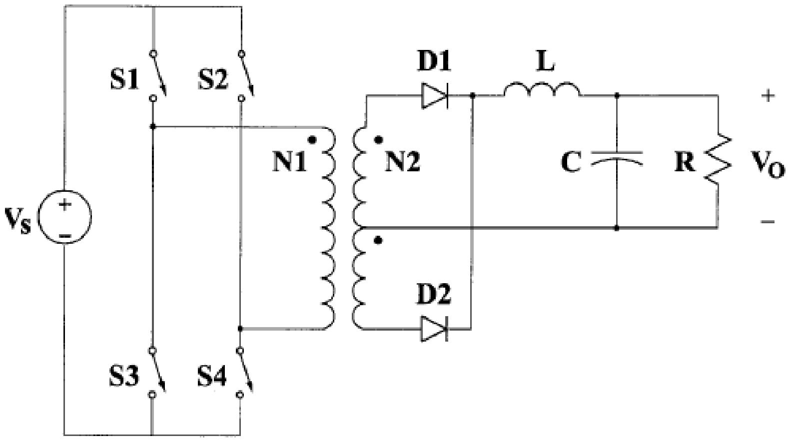


Figure 13.8 Full-bridge converter

Comparing the PWM dc-dc full-bridge converter of Fig. 13.8 to the half-bridge converter, it can be seen that the input capacitors have been replaced by two controllable switches that are operated in pairs. When S1 and S4 are on, voltage V_S is applied to the primary switch of the transformer and diode D1conducts. With S2 and S3 on, there is voltage $-V_S$ across the transformer primary switch

and diode D2 is on. With all controllable switches off, both diodes conduct in the same way as in the push-pull and half-bridge converters.
The dc voltage transfer function of the full-bridge converter is

$$M_V = \frac{V_O}{V_S} = \frac{2D}{n} \qquad (13.12)$$

where D ≤ 0.5
The values of filter components can be obtained from Eqs. (13.9) and (13.10).
It should be stressed that the full-bridge topology is a very versatile one. With different control algorithms, it is very popular in dc-ac conversion (square-wave and PWM single phase inverters), and it is also used in four-quadrant dc drives.

Step-Up DC/DC Converter (Boost Converter):

Figure 13.9a depicts a step-up or a PWM boost converter. It consists of dc input voltage source V_S, boost inductor L, controlled switch S, diode D, filter capacitor C, and load resistance R.

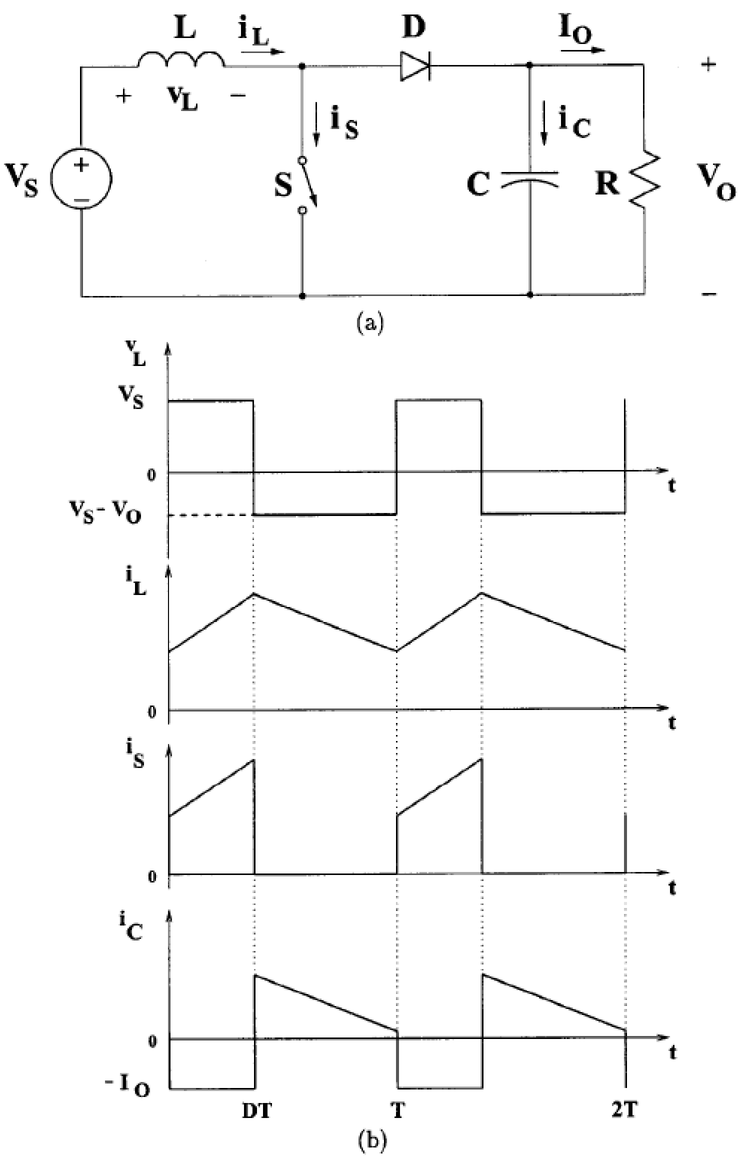


Figure 13.9 Boost converter: (a) circuit diagram; (b) waveforms.

The converter waveforms in the CCM are presented in Fig. 13.9b. When the switch S is in the on state, the current in the boost inductor increases linearly and the diode D is off at that time. When the switch S is turned off, the energy stored in the inductor is released through the diode to the output RC circuit.

Using Faraday's law for the boost inductor

$$V_S DT = (V_O - V_S)(1 - D)T \qquad (13.13)$$

from which the dc voltage transfer function turns out to be

$$M_V = \frac{V_O}{V_S} = \frac{1}{1 - D} \qquad (13.14)$$

As the name of the converter suggests, the output voltage is always greater than the input voltage. The boost converter operates in the CCM for L > Lb where

$$L_b = \frac{(1 - D)^2 DR}{2f} \qquad (13.15)$$

Ex. 4:

For D = 0.5, R = 10 Ω, and f = 100 kHz,
the boundary value of the inductance is L_b = 6.25 μH.

As shown in Fig. 13.9b, the current supplied to the output RC circuit is discontinuous. Thus, a larger filter capacitor is required in comparison to that in the buck-derived converters to limit the output voltage ripple. The filter capacitor must provide the output dc current to the load when the diode D is off. The minimum value of the filter capacitance that results in the voltage ripple Vr is given by

$$C_{\min} = \frac{DV_O}{V_r Rf} \qquad (13.16)$$

Ex. 4:

At D =0.5, V_r / V_O = 1%, R = 10 Ω, and f = 100 kHz,
The minimum capacitance for the boost converter is Cmin = 50 μF.

The boost converter does not have a popular transformer (isolated) version.

Buck-Boost Converter

Basic Converter

A nonisolated (transformerless) topology of the buck-boost converter is shown in Fig.13.10a. The converter consists of dc input voltage source V_S, controlled switch S, inductor L, diode D, filter capacitor C, and load resistance R.
With the switch on, the inductor current increases while the diode is maintained off.
When the switch is turned off, the diode provides a path for the inductor current.

Note the polarity of the diode that results in its current being drawn from the output.

The buck-boost converter waveforms are depicted in Fig.13.10b.
The condition of a zero volt-second product for the inductor in steady state yields

$$V_S DT = -V_O(1-D)T \qquad (13.17)$$

Hence, the dc voltage transfer function of the buck-boost converter is

$$M_V \equiv \frac{V_O}{V_S} = -\frac{D}{1-D} \qquad (13.18)$$

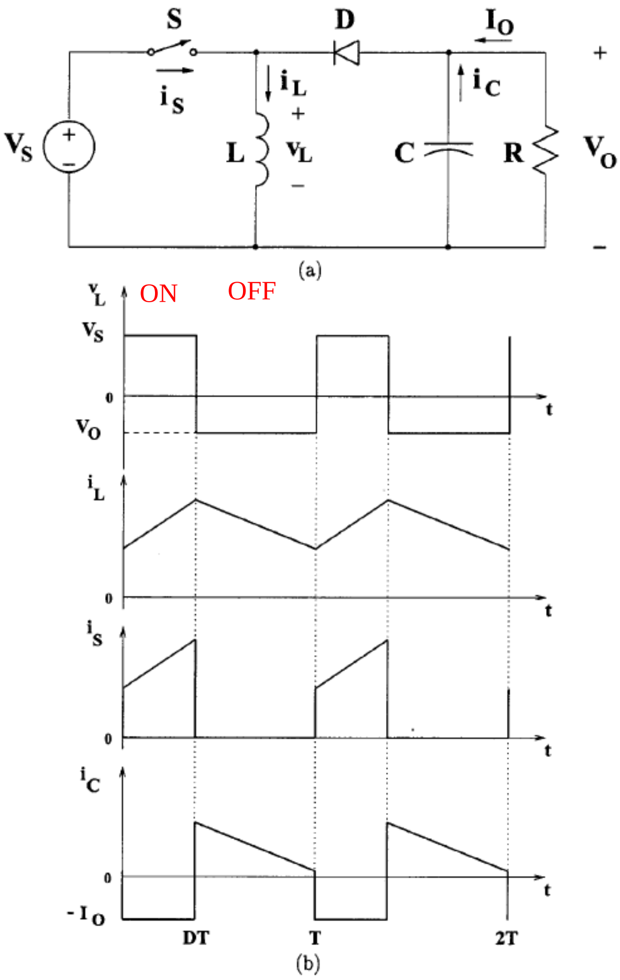


Figure 13.10 Buck-boost converter: (a) circuit diagram; (b) waveforms.

The output voltage V_O is negative with respect to the ground.
Its magnitude can be either greater or smaller (equal at D = 0.5) than the input voltage as the name of the converter implies.

The value of the inductor that determines the boundary between the CCM and DCM is

$$L_b = \frac{(1-D)^2 R}{2f} \qquad (13.19)$$

The structure of the output part of the converter is similar to that of the boost converter (reversed polarities are the only difference).
Thus, the value of the filter capacitor can be obtained from Eq.(13.16).

Flyback Converter

The **flyback converter** is used in both AC/DC and DC/DC conversion with galvanic isolation between the input and any outputs. The flyback converter is a buck-boost converter with the inductor split to form a transformer, so that the voltage ratios are multiplied with an additional advantage of isolation. When driving for example a plasma lamp or a voltage multiplier the rectifying diode of the boost converter is left out and the device is called a flyback transformer.

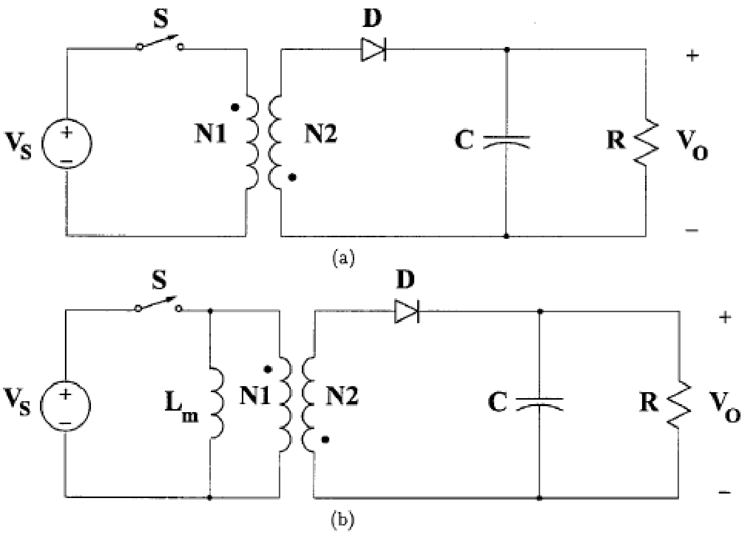
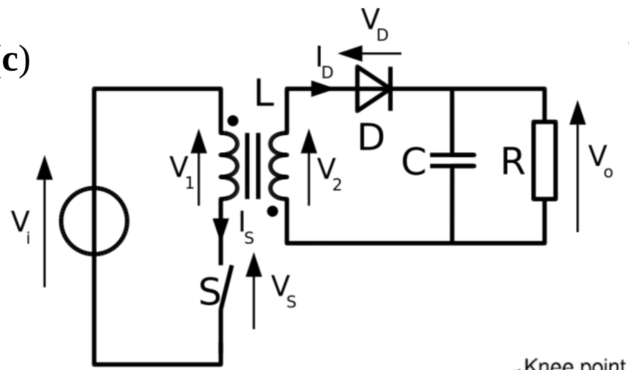
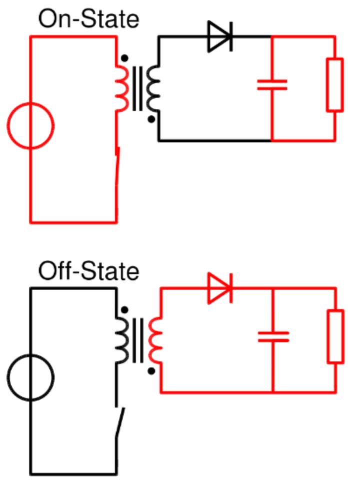


Figure 13.11 Flyback converter:
(a) circuit diagram;
(b) circuit with a transformere model showing the magnetizing inductance

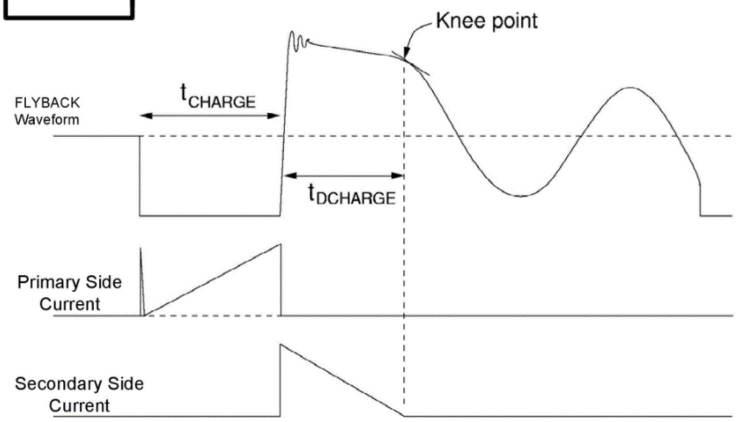
(c)





(d)

(e)



(c) Schematic of a flyback converter
(d) The two configurations of a flyback converter in operation: In the on-state, the energy is transferred from the input voltage source to the transformer (the output capacitor supplies energy to the output load). In the off-state, the energy is transferred from the transformer to the output load (and the output capacitor).
(e) Waveform - using primary side sensing techniques - showing the 'knee point'.
The schematic of a flyback converter can be seen in Fig. c. It is equivalent to that of a buck-boost converter,[1] with the inductor split to form a transformer. Therefore the operating principle of both converters is very close:

- When the switch is closed (top of Fig. d), the primary of the transformer is directly connected to the input voltage source. The primary current and magnetic flux in the transformer increases, storing energy in the transformer. The voltage induced in the secondary winding is negative, so the diode is reverse-biased (i.e., blocked). The output capacitor supplies energy to the output load.

- When the switch is opened (bottom of Fig. d), the primary current and magnetic flux drops. The secondary voltage is positive, forward-biasing the diode, allowing current to flow from the transformer. The energy from the transformer core recharges the capacitor and supplies the load.

The operation of storing energy in the transformer before transferring to the output of the converter allows the topology to easily generate multiple outputs with little additional circuitry, although the output voltages have to be able to match each other through the turns ratio. Also there is a need for a controlling rail which has to be loaded before load is applied to the uncontrolled rails, this is to allow the PWM to open up and supply enough energy to the transformer.

Operation

The flyback converter is an isolated power converter. The two prevailing control schemes are voltage mode control and current mode control (in the majority of cases current mode control needs to be dominant for stability during operation). Both require a signal related to the output voltage. There are three common ways to generate this voltage. The first is to use an optocoupler on the secondary circuitry to send a signal to the controller. The second is to wind a separate winding on the coil and rely on the cross regulation of the design. The third consists on sampling the voltage amplitude on the primary side, during the discharge, referenced to the standing primary DC voltage.

The first technique involving an optocoupler has been used to obtain tight voltage and current regulation, whereas the second approach has been developed for cost-sensitive applications where the output does not need to be as tightly controlled, but up to 11 components including the optocoupler could be eliminated from the overall design. Also, in applications where reliability is critical, optocouplers can be detrimental to the MTBF (Mean Time Between Failure) calculations. The third technique, primary-side sensing, can be as accurate as the first and more economical than the second, yet requires a minimum load so that the discharge-event keeps occurring, providing the opportunities to sample the 1:N secondary voltage at the primary winding (during Tdischarge, as per Fige).

A variation in primary-side sensing technology is where the output voltage and current are regulated by monitoring the waveforms in the auxiliary winding used to power the control IC itself, which have improved the accuracy of both voltage and current regulation. The auxiliary primary winding is used in the same discharge phase as the remaining secondaries, but it builds a rectified voltage referenced commonly with the primary DC, hence considered on the primary side.

Previously, a measurement was taken across the whole of the flyback waveform which led to error, but it was realized that measurements at the so-called *knee point* (when the secondary current is zero, see Fig. e) allow for a much more accurate measurement of what is happening on the secondary side. This topology is now replacing ringing choke converters (RCCs) in applications such as mobile phone chargers.

Limitations

Continuous mode has the following disadvantages, which complicate the control of the converter:

- The voltage feedback loop requires a lower bandwidth due to a right half plane zero in the response of the converter.
- The current feedback loop used in current mode control needs slope compensation in cases where the duty cycle is above 50%.
- The power switches are now turning on with positive current flow - this means that in addition to turn-off speed, the switch turn-on speed is also important for efficiency and reducing waste heat in the switching element.

Discontinuous mode has the following disadvantages, which limit the efficiency of the converter:

1- High RMS and peak currents in the design
2- High flux excursions in the inductor

Applications

1. 1-Low-power switch-mode power supplies (cell phone charger, standby power supply in PCs)
2. 2-Low-cost multiple-output power supplies (e.g., main PC supplies <250W[citation needed])
3. 3-High voltage supply for the CRT in TVs and monitors (the flyback converter is often combined with the horizontal deflection drive)
4. 4-High voltage generation (e.g., for xenon flash lamps, lasers, copiers, etc.)
5. 5-Isolated gate driver

Best Answer: The best explanation I've ever heard is that it was named from the similar circuit the "flyback transformer". This is what is used to provide the low-power high-voltage needed to drive a CRT. The name comes from this use where a sawtooth wave drives the transformer to build up a voltage to draw an electron beam across the display. Each time it switches off, it causes the e-beam to "fly back" from the right to the left of the screen.

Galvanic isolation is a principle of isolating functional sections of electrical systems to prevent current flow; no direct conduction path is permitted.[1] Energy or information can still be exchanged between the sections by other means, such as capacitance, induction or electromagnetic waves, or by optical, acoustic or mechanical means.

A PWM flyback converter is a very practical isolated version of the buck-boost converter. The circuit of the flyback converter is presented in Fig.13.11a. The inductor of the buck-boost converter has been replaced by a flyback transformer.
The input dc source V_S and switch S are connected in series with the transformer primary. The diode D and the RC output circuit are connected in series with the secondary of the flyback transformer. Figure13.11b shows the converter with a simple flyback transformer model that includes a magnetizing inductance Lm and an ideal transformer with a turns ratio

n = N1/N2.

The flyback transformer leakage inductances and losses are neglected in the model. It should be noted that leakage inductances, although not important from the viewpoint of the principle of operation, affect adversely switch and diode transitions. Therefore, snubbers are usually required in flyback converters.

Refer to Fig.13.11b for the converter operation. When the switch S is on, the current in the magnetizing inductance increases linearly, the diode D is off and there is no current in the ideal transformer windings. When the switch is turned off, the magnetizing inductance current is diverted into the ideal transformer, the diode turns on, and the transformed magnetizing inductance current is supplied to the RC load.
The dc voltage transfer function of the flyback converter is

$$M_V = \frac{V_O}{V_S} = \frac{D}{n(1-D)} \qquad (13.20)$$

It differs from the buck-boost converter voltage transfer function by the turns ratio factor n. A positive sign has been obtained by an appropriate coupling of the transformer windings.
Unlike in transformer buck-derived converters, the magnetizing inductance Lm of the flyback transformer is an important design parameter. The value of the magnetizing inductance that determines the boundary between the CCM and DCM is given by

$$L_{mb} = \frac{n^2(1-D)^2 R}{2f} \qquad (13.21)$$

The value of the filter capacitance can be calculated using Eq. (13.16).

http://www.dos4ever.com/flyback/flyback.html
http://en.wikipedia.org/wiki/Flyback_converter
[1] http://www.williamson-labs.com/555-tutorial.htm
[2] http://www.uoguelph.ca/~antoon/gadgets/555/555.html
https://answers.yahoo.com/question/index;_ylt=A0LEVzvuU2hT2RkABQxXNyoA;_ylu=X3oDMTEzNWpuM3RhBHNlYwNzcgRwb3MDMwRjb2xvA2JmMQR2dGlkA1ZJUDM5OV8x?qid=20070628222838AAbeZru

C`uk Converter

The circuit of the C`uk converter is shown in Fig.13.12a. It consists of

1. dc input voltage source V_S,
2. input inductor L1,
3. controllable switch S,
4. energy transfer capacitor C1,
5. diode D,
6. filter inductor L2,
7. filter capacitor C,
8. and load resistance R.

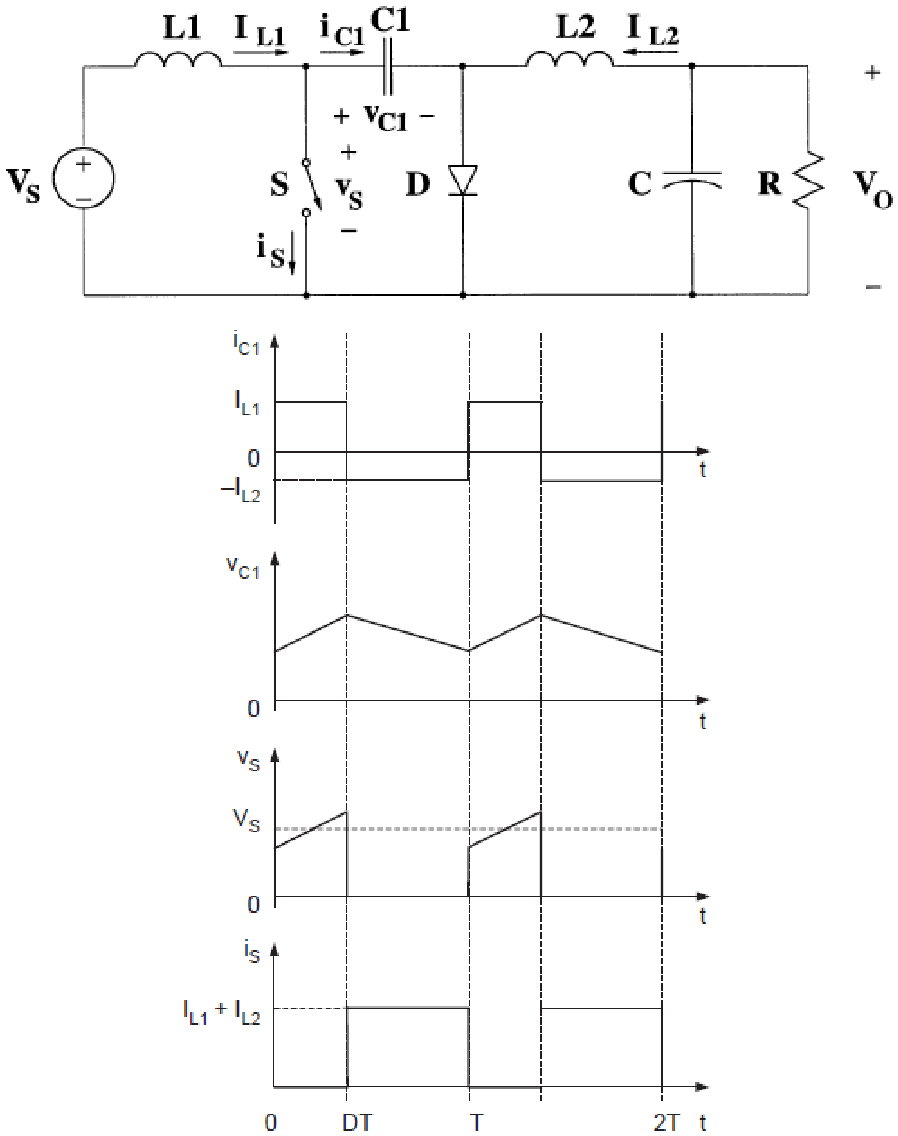


Figure 13.12 C`uk converter:
(a) circuit diagram;
(b) waveforms.

- An important advantage of this topology is a **continuous current** at both the input and the output of the converter.
- Disadvantages of the C`uk converter are a high number of reactive components **and** high current stresses on the switch, the diode, and the capacitor C1.

The main waveforms in the converter are presented in Fig. 13.12b. When the switch is on, the diode is off and the capacitor C1 is discharged by the inductor L2 current. With the switch in the off state, the diode conducts currents of the inductors L1 and L2, whereas capacitor C1 is charged by the inductor L1 current.

To obtain the dc voltage transfer function of the converter, we shall use the principle that the average current through a capacitor is zero for steady-state operation. Let us assume that inductors L1 and L2 are large enough that their ripple current can be neglected. Capacitor C1 is in steady state if

$$I_{L2} DT = I_{L1}(1 - D)T \qquad (13.22)$$

For a lossless converter

$$P_S = V_S I_{L1} = -V_O I_{L2} = P_O \qquad (13.23)$$

Combining these two equations, the dc voltage transfer function of the C`uk converter is

$$M_V \equiv \frac{V_O}{V_S} = -\frac{D}{1-D} \qquad (13.24)$$

This voltage transfer function is the same as that for the buckboost converter.
The boundaries between the CCM and DCM are determined by

$$L_{b1} = \frac{(1-D)^2 R}{2Df} \qquad (13.25)$$

for L1 and

$$L_{b2} = \frac{(1-D)R}{2f} \qquad (13.26)$$

for L2.

The output part of the C`uk converter is similar to that of the buck converter.
Hence, the expression for the filter capacitor C is

$$C_{\min} = \frac{(1-D)V_O}{8V_r L_2 f^2} \qquad (13.27)$$

The peak-to-peak ripple voltage in the capacitor C1 can be estimated as

$$V_{r1} = \frac{DV_O}{C_1 R_f} \qquad (13.28)$$

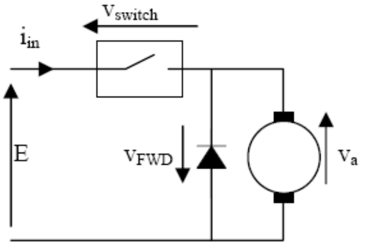
A transformer (isolated) version of the C`uk converter can be obtained by splitting capacitor C1 and inserting a high frequency transformer (Pulse Transformer) between the split capacitors.

Example 5:

A buck dc/dc converter is used to drive a dc machine. The frequency is such that the armature current may be considered to be constant and its value is 5 A. The average armature voltage is 25 V. The input voltage to the converter is 100 V. The semiconductor switch voltage drop is 5 V and the FW diode drop is 2 V.
If the cycle time is T, sketch the armature voltage and switch current waveforms. Compute the value of time ratio and the average value of the input current to the converter.

Solution:

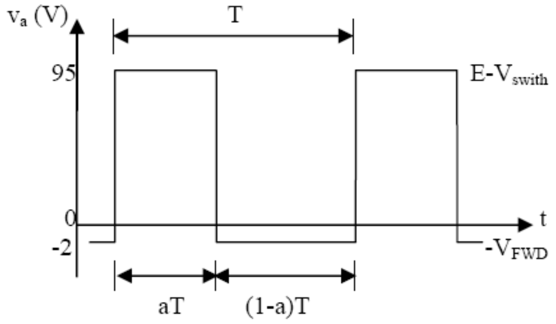
A circuit diagram is drawn with self-explanatory labels. Using the armature voltage waveform, the average armature voltage is given by:



$$V_a = \frac{1}{T}\int_{t=t_o}^{t_o+T} v_a dt$$

$$V_a = \frac{(E - V_{switch})\ T_{ON} - V_{FWD} T_{OFF}}{T}$$

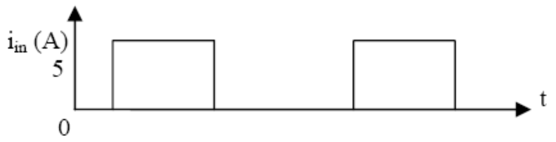
$$V_a = (E - V_{switch})\ a - V_{FWD}\ (1-a)$$



Therefore,

$$a = \frac{V_a + V_{FWD}}{(E - V_{switch}) + V_{FWD}}$$

$$a = \frac{25+2}{97} = 0.2784$$



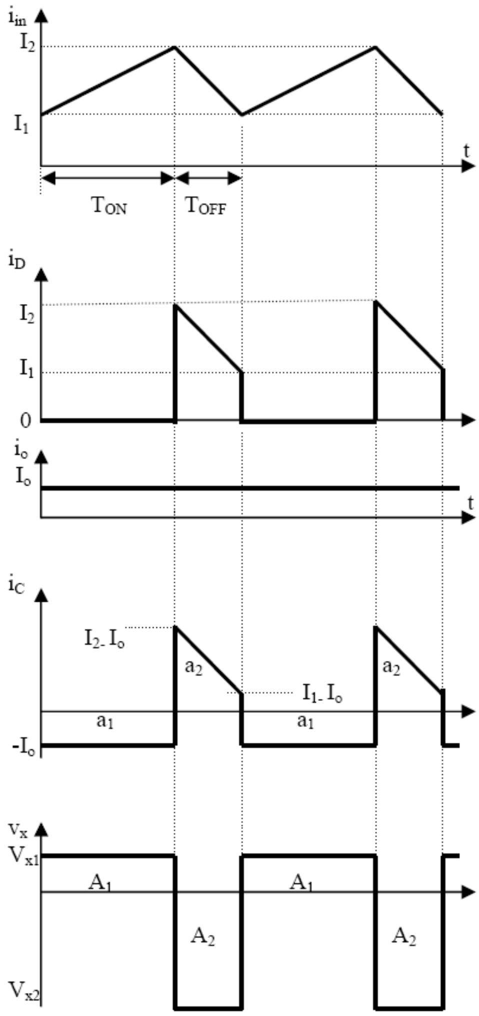
The average value of the input current is:

$$I_{in} = a\ I_{av} = 0.2784 \text{ x } 5 = 1.392 \text{ A}$$

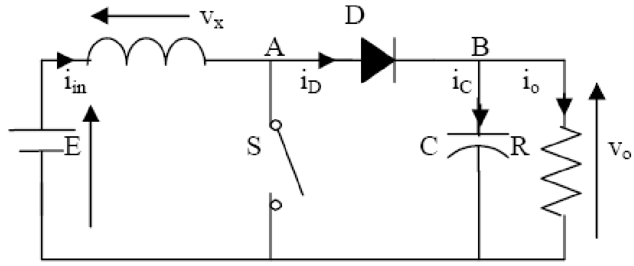
Boost Converters:

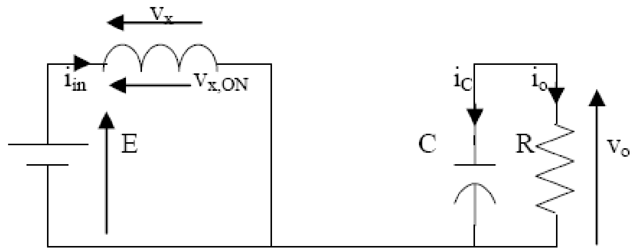
In contrast to the step-down chopper, the minimum average output voltage of the step-up chopper equals the input voltage "E". Again, changing the duty cycle can change the average output voltage (V_o).

The principle of operation of this chopper is the storage of energy at low supply voltage (in an inductor) and its release at higher voltage. A transistor is used to act as a switch for transferring the energy in the storage component (the inductor) to the output circuit.

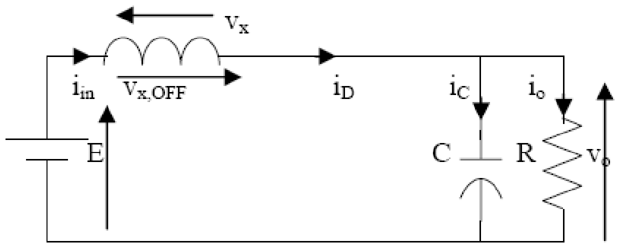


V_{x1}= E and V_{x2} = - (Vo - E)
Area a_1 = a_2: they represent, respectively, charge transferred from C to load and from source to C, and consequently electrical energy.
Area A_1 = A_2: they represent when multiplied by input current, stored and released magnetic energy, in the inductor L, respectively.



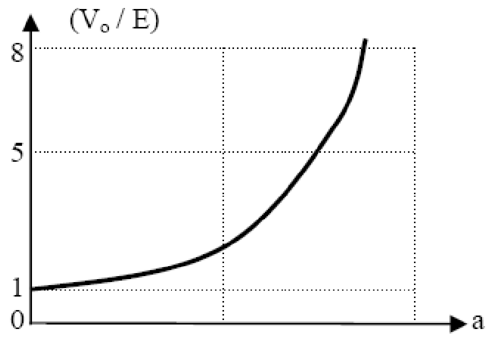


ON-State Equivalent Circuit



OFF-State Equivalent Circuit

The voltage regulation of practical chopper is very poor. A feedback regulator is therefore very essential for operation with variable load to stabilize the output voltage.



Ex.5:

A boost converter is used to convert 6 V input to supply 24 W at an output voltage of 48V±0.5 %. The input current variation shall be no more than ± 1%. If the switching frequency is 20 kHz, specify the smallest inductance and capacitance to meet the given specifications.

Solution:

$$V_o = \frac{E}{1-a}; a = 1 - \frac{E}{V_o} = 1 - \frac{6}{48} = 0.875$$

$$T = \frac{1}{f} = \frac{1}{20kHz} = 50\mu s \qquad T_{ON} = aT = \ldots = 43.75\ \mu s, \text{ and } T_{OFF} = (1-a)T = 6.25\ \mu s$$

$$P_o = I_o V_o; \qquad I_o = 24/48 = 0.5A$$

Assuming the converter is operating with low losses, then

$$P_o \cong P_{in} = V_{in} I_{in}; \qquad I_{in} = 24/6 = 4\ A$$

During the ON period of the chopper $E = L\frac{di_{in}}{dt} \cong L\frac{\Delta I_{in}}{T_{ON}}$

The maximum allowable variation in the input current $(\Delta I_{in} / I_{in})$ is $\pm 1\%$. The peak-to-peak variation of the input current is therefore 2% and hence

$$\Delta I_{in} = \frac{2}{100} I_{in} = \frac{2}{100}(4) = 0.08\ A$$

Since $E = L\frac{\Delta I_{in}}{T_{ON}}$, the required inductance is determined from $L > \frac{ET_{ON}}{\Delta I_{in}}$. Therefore

$$L \geq \frac{6x43.75x10^{-6}}{0.08}, \qquad \text{or } L \geq 3.28\ mH$$

The capacitor has to supply an average current of 0.5A for the entire ON period "T_{ON}" such that V_o will not change by more than ± 0.5 % of the rated value. Thus the peak-to-peak variation of the output voltage is

$$\Delta V_o = \frac{1}{100} V_o = \frac{1}{100}(48) = 0.48V$$

This variation in the output voltage is approximately determined from

$$I_o = I_c = C\frac{dv_c}{dt} = C\frac{dv_o}{dt} \cong C\frac{\Delta V_o}{T_{ON}}$$

Thus,

$$\Delta V_o = \frac{I_o T_{ON}}{C},$$

$$C \geq \frac{I_o T_{ON}}{\Delta V_o} \qquad \text{or } C \geq \frac{0.5x43.75\,x10^{-6}}{0.48}$$

The required capacitance is at least 45.6 μF.

Ex.6:

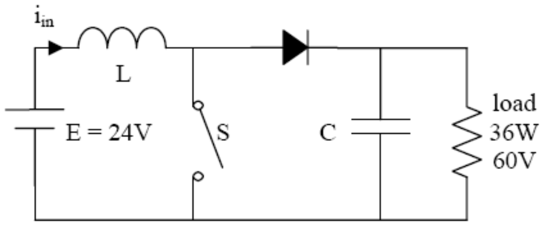
The switching frequency of the dc-dc converter shown is 20 kHz.

(A). Calculate a, Iin and draw the equivalent circuit at source terminals.

(B). Find the minimum value of "L" if the maximum variation of input current should be less than ± 2%.

(C). Find the minimum value of "C" if the minimum value of the output voltage is 59.4V.

Solution



(A). This is a step-up converter since it boosting up the input voltage (24V) up to 60V at the output.

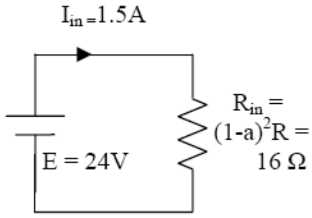
$$V_o = \frac{E}{1-a};\ a = 1-\frac{E}{V_o} = 1-\frac{24}{60} = 0.6$$

$$R = \frac{V_o^2}{P_o} = \frac{60^2}{36} = 100\ \Omega\ ; \qquad I_o = Po / Vo = 36 / 60 = 0.6A$$

$$I_{in} = \frac{I_o}{1-a} = \frac{0.6}{1-0.6} = 1.5\ A$$

$$R_{in} = \frac{E}{I_{in}} = \frac{(1-a)V_o}{I_o/(1-a)} = (1-a)^2 R = (1-0.6)^2(100) = 16\ \Omega$$

The equivalent circuit at source terminals is as shown in the adjacent diagram.



(B). The maximum allowable variation in the input current is $\pm 2\%$. The peak-to-peak variation of the input current is therefore 4% and hence

$$\Delta I_{in} = 4\% \text{ of } I_{in} = 0.04\ (1.5) = 0.06\ A$$

$$T = \frac{1}{f} = \frac{1}{20kHz} = 50\mu s$$

T_{ON} = aT = 0.6 x 50 = 30 μs, and T_{OFF} = (1- a)T = 04 x 50 = 20 μs

During the ON period of the converter $E = L\frac{di_{in}}{dt} \cong L\frac{\Delta I_{in}}{T_{ON}}$

Since $E = L\frac{\Delta I_{in}}{T_{ON}}$, the required inductance is determined from $L \geq \frac{ET_{ON}}{\Delta I_{in}}$. Therefore

$$L \geq \frac{24\ x\ 30\ x\ 10^{-6}}{0.06} \qquad \text{or } L \geq 12\text{ mH.}$$

Alternatively, one may use the OFF period data to find "L" as follows. During the OFF period:

$$E = L\frac{di_{in}}{dt} + R\ i_o \cong L\frac{\Delta I_{in}}{T_{OFF}} + RI_o$$

$$24 = L\frac{-0.06}{20\ x\ 10^{-6}} + 60$$

$$L \geq \frac{36\ x\ 20\ x\ 10^{-6}}{0.06} \qquad \text{or } L \geq 12\text{ mH.}$$

(C). The capacitor has to supply an average current of 0.6 A for the entire ON period such that V_o will not drop below 59.4 V. The average output voltage is 60V and therefore the output voltage is varying between 59.4 V and 60.6 V giving a peak-to-peak variation (ΔV_o) of 1.2V.

This variation in the output voltage is approximately determined from

$$I_o = I_c = C\ \frac{dv_c}{dt} = C\frac{dv_o}{dt} \cong C\frac{\Delta V_o}{T_{ON}}$$

Thus,

$$\Delta V_o = \frac{I_o T_{ON}}{C},$$

$$C > \frac{I_o T_{ON}}{\Delta V_o} \qquad \text{or } C > \frac{0.6\ x\ 30\ x\ 10^{-6}}{1.2}$$

The required capacitance is at least 15 μF.

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