Chapter 6

Soft-Switching dc-dc Converters

A new class of dc-to-dc converters, known in the literature as *soft-switching resonant converters,* has been thoroughly investigated in recent years. Soft switching means that one or more power switches in a dc-dc converter have either the turn-on or turn-off switching losses eliminated. This is in contrast to hard switching, where both turn-on and turn-off of the power switches are done at high current and high voltage levels. One approach is to create a full-resonance phenomenon within the converter through series or parallel combinations of resonant components. Such converters are generally known as resonant converters. Another approach is to use a conventional PWM converter—buck, boost, buck-boost, Cuk, SEPIC—and replace the switch with a resonant switch that accomplishes the loss elimination. Because of the nature of the PWM circuit, resonance occurs for a shorter time interval compared to the full-resonance case. This class of converters, combining resonance and PWM, is appropriately known as quasi-resonance converters. In this chapter, our focus will be on the latter method, mainly using the resonance PWM switch to achieve soft switching. For simplicity, here we use the term soft switching to refer to dc-dc converters, quasi-resonance converters, and other topologies that employ resonance to reduce switching losses. Two major techniques are employed to achieve soft switching: zero-current switching (ZCS) and zero-voltage switching (ZVS). This chapter will focus on ZCS and ZVS types of PWM resonant switches and their steady-state analyses.

6.1 TYPES OF DC-DC CONVERTERS

As shown in previous chapters, linear-mode and switch-mode converters have been used widely in the design of commercial dc-to-dc power supplies. Linear power supplies offer the designer four major advantages: simplicity in design, no electrical noise in the output, fast dynamic response time, and low cost. Their applications, however, are limited due to several disadvantages: (1) The input voltage is at least 2 or 3 V higher than the output voltage because the circuit can only be used as a step-down regulator, (2) each regulator is limited to only one output, and (3) efficiency is low compared to other switching regulators (30% to 60% for an output voltage less than 20 V).

High-frequency pulse-width-modulation (PWM) switching regulators overcome all the linear regulators' shortcomings: (1) They have higher efficiency ($>$90%); (2) power transistors operate at their most efficient points—cutoff and saturation—allowing for power densities of around 40 to 50 W/in^3; (3) multi-output applications are possible; and (4) the size and the cost are much lower, especially at high power levels.

However, PWM switching converters still have several limitations, among them (1) greater circuit complexity compared to the linear power supplies, (2) high electromagnetic interference (EMI), and (3) switching speeds below 100 kHz because of high stress levels on power semiconductor devices.

A third generation of power converters was introduced in the late 1980s. This group of topologies is known as soft-switching resonant converters. Compared with linear and switched-mode converters, the potential advantage of the soft-switching resonant converters is reduced power losses, thus achieving high switching frequency and high power density while maintaining high efficiency. Moreover, due to the higher switching frequency, such converters exhibit faster transient responses. Today's soft-switching techniques are used in the design of both high-frequency dc-to-dc conversion and high-frequency dc-to-ac inversion. Only the first application is discussed in this text.

The Resonance Concept

Like switched-mode dc-to-dc converters, resonant converters are used to convert dc to dc through an additional stage: the resonant stage, in which the dc signal is converted to a high-frequency ac signal. The advantages of the resonant converter include the natural commutation of power switches, resulting in low switching power dissipation and reduced component stresses, which in turn results in increased power efficiency and increased switching frequency; and higher operating frequencies, resulting in reduced size and weight of equipment and in faster response, and hence a possible reduction in EMI problems.

Since the size and weight of the magnetic components (inductors and transformers) and the capacitors in a converter are inversely proportional to the converter's switching frequency, many power converters have been designed for

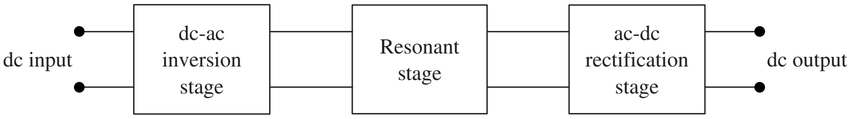

Figure 6.1 Typical block diagram of soft-switching dc-to-dc converter.

progressively higher frequencies in order to reduce the size and weight and to obtain fast converter transients. In recent years, the market demand for wide applications that require variable-speed drives, highly regulated power supplies, and uninterruptible power supplies, as well as the desire for smaller size and lighter weight, has increased.

There are many soft-switching techniques available in the literature to improve the switching behavior of dc-to-dc resonant converters. At the time of this writing, intensive research in soft switching is under way to further improve the efficiency through increased switching frequency of power electronic circuits.

From a circuit standpoint, a dc-to-dc resonant converter can be described by three major circuit blocks as shown in Fig. 6.1: the dc-to-ac input inversion circuit, the resonant energy buffer tank circuit, and the ac-to-dc output rectifying circuit. Typically, the dc-to-ac inversion is achieved by using various types of switching network topologies. The resonant tank, which serves as an energy buffer between the input and the output, is normally synthesized by using a lossless frequency-selective network. The purpose of that network is to regulate the energy flow from the source to the load. Finally, the ac-to-dc conversion is achieved by incorporating rectifier circuits at the output section of the converter.

Resonant versus Conventional PWM

For many years, high-efficiency power-processing circuits have been achieved by operating power semiconductor devices in the switching mode, whereby switching devices are operated in either the on or off state, as in the PWM method. In PWM converters, the switching of semiconductor devices normally occurs at high current levels. Therefore, when switching at high frequencies these converters are associated with high power dissipation in their switching devices. Furthermore, PWM converters suffer from EMI caused by high-frequency harmonic components associated with their quasi-square switching current and/or voltage waveforms. Unfortunately, even though the technological advancements of PWM switch-mode converters has resulted in faster switching, their operating frequency is limited by the factors mentioned above.

In the resonant technique, switching losses in the semiconductor devices are avoided due to the fact that the current through or voltage across the switching device at the switching point is equal to or near zero. This reduction in switching losses allows the designer to attain a higher operating frequency without sacrificing the converter's efficiency. Compared to the PWM converters, the resonant converters show the promise of achieving the design of small-size and low-weight converters. Currently, resonant power converters operating in the range of a few megahertz are available. Another advantage of resonant converters over PWM converters is the decrease in the harmonic content in the converter voltage and current waveforms. Therefore, when the resonant and PWM converters are operated at the same power level and frequency, the resonant converter can be expected to have lower harmonic emission.

6.2 CLASSIFICATION OF SOFT-SWITCHING RESONANT CONVERTERS

The literature is very rich with resonant power electronic circuits used in applications such as dc-to-dc and dc-to-ac resonant converters. To date, there exists no general classification of resonant converter topologies. In fact, more and more resonant topologies continue to be introduced in the open literature. Several types of dc-to-dc converters employing additional resonant stages that have been explored thus far may be summarized as follows:

- Quasi-resonant converters (single-ended)
  - Zero-current switching (ZCS)
  - Zero-voltage switching (ZVS)
- Full-resonance converters (conventional)
  - Series resonant converter (SRC)
  - Parallel resonant converter (PRC)
- Quasi-squarewave (QSW) converters
  - Zero-current switching (ZCS)
  - Zero-voltage switching (ZVS)
- Zero-clamped topologies
  - Zero-clamped-voltage (CV)
  - Zero-clamped-current (CC)
- Class E resonant converters
- dc link resonant inverters
- Multi-resonant converters
  - Zero-current switching (ZCS)
  - Zero-voltage switching (ZVS)
- Zero Transition Topologies
  - Zero-voltage transition (ZVT).
  - Zero-current transition (ZCT).

Many other variations of soft-switching topologies that exist today are beyond the scope of this text. Since this text targets senior undergraduate electrical engineering students, we focus mainly on the quasi-resonant type of PWM converters.

6.3 ADVANTAGES AND DISADVANTAGES OF ZCS AND ZVS

The major advantage of ZCS and ZVS quasi-resonant converters is that the power switch is turned on and off at zero voltage and zero current, respectively. In ZCS topologies the rectifying diode has ZVS, whereas in ZVS topologies the rectifying diode has ZCS. A second advantage is that both ZVS and ZCS converters utilize transformer leakage inductors and diode junction capacitors and the output parasitic capacitor of the power switch.

The major disadvantage of the ZVS and ZCS techniques is that they require variable-frequency control to regulate the output. This is undesirable since it complicates the control circuit and generates unwanted EMI harmonics, especially under wide load variations. In ZCS, the power switch turns off at zero current, but at turn-on the converter still suffers from the capacitor turn-on loss caused by the output capacitor of the power switch.

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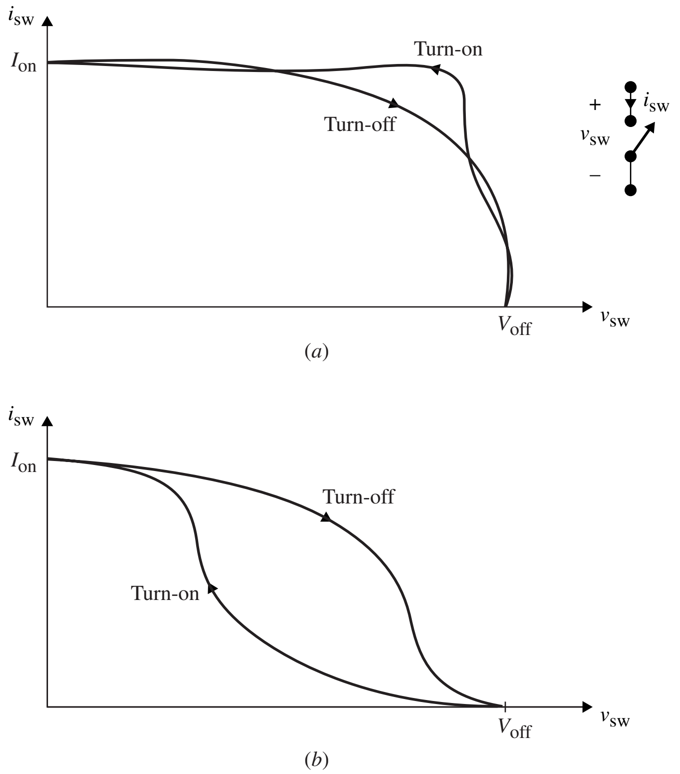

Figure 6.2 Switching loci. (*a*) Without snubber circuit. (*b*) With snubber circuit.

Switching Loci

Most regulator converter switches need to turn on or turn off the full load current at a high voltage, resulting in what is known as hard switching. Figure 6.2(*a*) and (*b*) shows typical switching loci for a hard-switching converter without and with a snubber circuit, respectively.

In a soft-switching converter topology, an *LC* resonant network is added to shape the switching device's voltage or current waveform into a quasi-sinewave in such a way that a zero voltage or a zero current condition is created. This technique eliminates the turn-on or turn-off loss associated with the charging or discharging of the energy stored in the MOSFET's parasitic junction capacitors. Figure 6.3(*a*) and (*b*) shows typical switching loci for the ZVS at turn-on and the ZCS at turn-off cases.

Switching Losses

As the frequency of operation increases, the switching losses also increase. As shown in Chapter 3, there are two types of switching losses:

1. At turn-off, the power transformer leakage inductance produces high di/dt, which results in a high voltage spike across it.
2. At turn-on, the switching loss is mainly caused by the dissipation of energy stored in the output parasitic capacitor of the power switch.

Figure 6.4(*a*) and (*b*) shows typical switching waveforms at turn-off and turn-on, respectively.

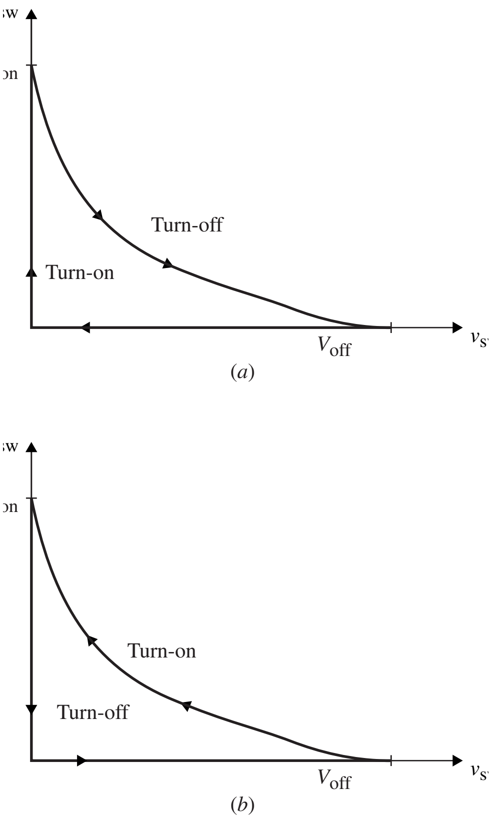

Figure 6.3 (*a*) ZVS at turn-on. (*b*) ZCS at turn-off.

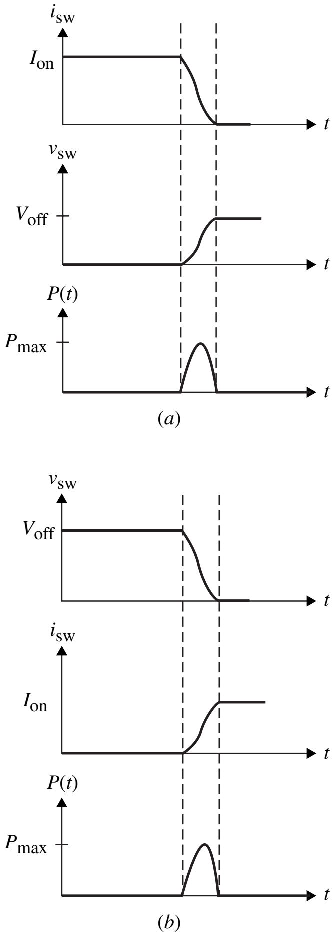

Figure 6.4 Typical switching current, voltage, and power loss waveforms at (*a*) turn-off and (*b*) turn-on.

6.4 ZERO-CURRENT SWITCHING TOPOLOGIES

6.4.1 The Resonant Switch

In this section, we will present one class of PWM converters that was introduced in the late 1980s based on the concept of using conventional PWM switching along with an *LC* tank circuit. Depending on the inductor-capacitor arrangements, there are two possible resonant switch configurations. The switch is either L-type or an M-type[1] and can be implemented as half-wave or full-wave, which corresponds to whether the switch current is unidirectional or bidirectional, respectively. Figure 6.5 shows the L-type switch in both the half- and full-wave implementations. The M-type switch is shown in Fig. 6.6.

The three conventional converter topologies–buck, boost, and buck-boost—will be analyzed here. In all of these topologies, the *LC* tank forms the resonant tank that causes ZCS to occur. The buck, boost, and buck-boost converters are shown in Fig. 6.7(*a*), (*b*), and (*c*), respectively.

[1]The notations *L-type* and *M-type* were used by the original authors.

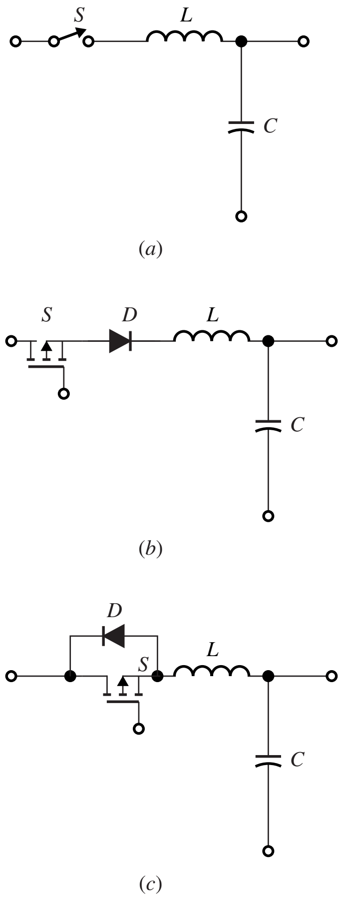

Figure 6.5 Resonant switch. (*a*) L-type switch. (*b*) Half-wave implementation. (*c*) Full-wave implementation.

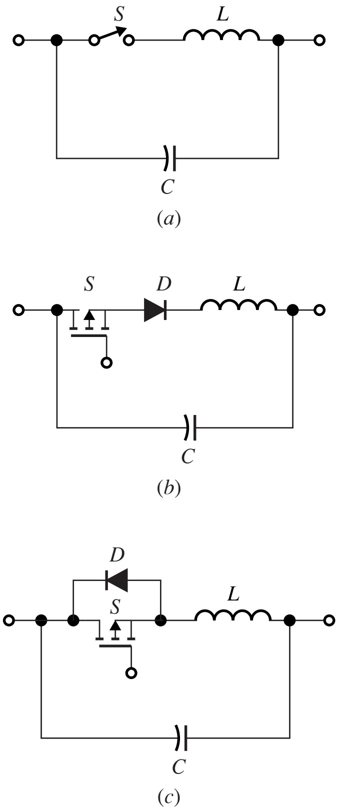

Figure 6.6 Resonant switch. (*a*) M-type switch. (*b*) Half-wave implementation. (*c*) Full-wave implementation.

The detailed steady-state analyses of these conventional converters was presented in Chapter 4. There exist two ccm modes of operation (as far as energy transfer is concerned): one mode during which energy is transferred from the source to the storage inductor, and a second mode during which energy is transferred from the storage inductor to the load. A low-pass LC filter is used in all three topologies to filter the fundamental frequency from the harmonics. Using the two types of switch arrangements, it is possible to convert the three topologies into quasi-resonant converters.

6.4.2 Steady-State Analysis of Quasi-Resonant Converters

To simplify the steady-state analysis of the converters, some assumptions need to be made:

1. The filtering components L_o, L_{in}, L_F, and C_o are very large compared to the resonant components L and C.
2. The output filter L_o-C_o-R is treated as a constant current source, I_o.
3. The output filter C_o-R is treated as a constant voltage source, V_o.
4. Switching devices and diodes are ideal.
5. Reactive circuit components are ideal.

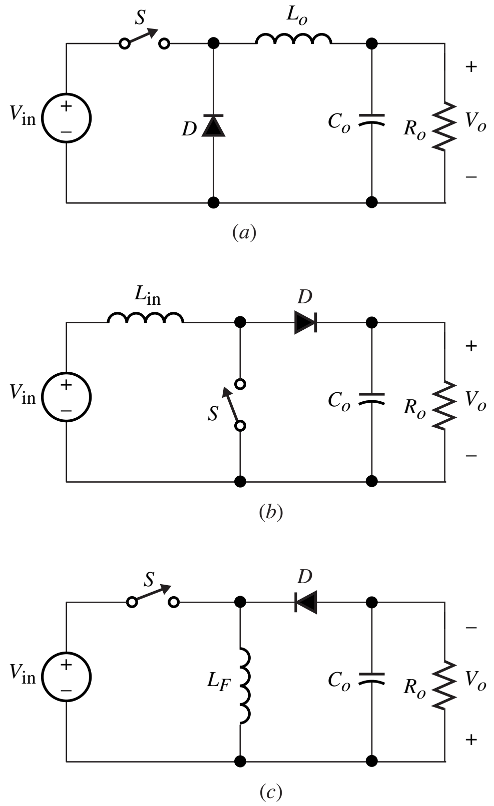

Figure 6.7 Conventional converters: (*a*) buck, (*b*) boost, and (*c*) buck-boost.

The Buck Resonant Converter

Replacing the switch in Fig. 6.7(*a*) by the resonant-type switch of Fig. 6.5(*a*), we obtain a quasi-resonant PWM buck converter, as shown in Fig. 6.8(*a*). Its simplified equivalent circuit is shown in Fig. 6.8(*b*). It can be shown that there are four modes of operation under the steady-state condition.

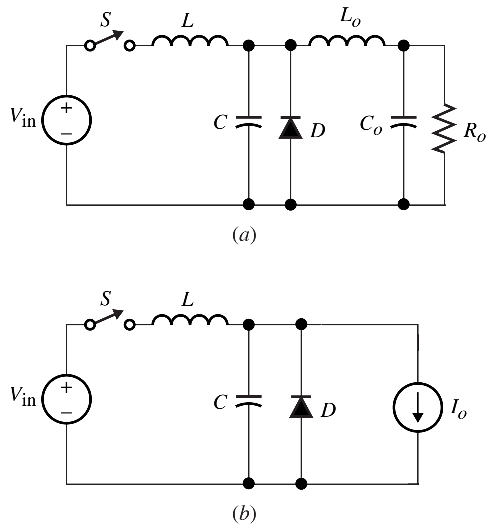

Figure 6.8 (*a*) Conventional buck converter with L-type resonant switch. (*b*) Simplified equivalent circuit.

Mode I **($0 \le t < t_1$)** Mode I starts at $t = 0$, when S is turned on. Since the switch was off prior to $t = 0$, it is clear that the diode must have been on for $t < 0$ to carry the output inductor current. Hence, we assume for $t > 0$, both S and D are on. The output current is equal to the constant current source, I_o, as shown in Fig. 6.9(*a*). In this mode, the capacitor voltage, v_c, is zero and the input voltage is equal to the inductor voltage as given by

$$V_{\text{in}} = L\frac{di_L}{dt} \tag{6.1}$$

Integrating Eq. (6.1) from 0 to t, the inductor current, i_L, is given by

$$i_L(t) = \frac{V_{\text{in}}}{L}t \tag{6.2}$$

Equation (6.2) assumes a zero initial condition for i_L.

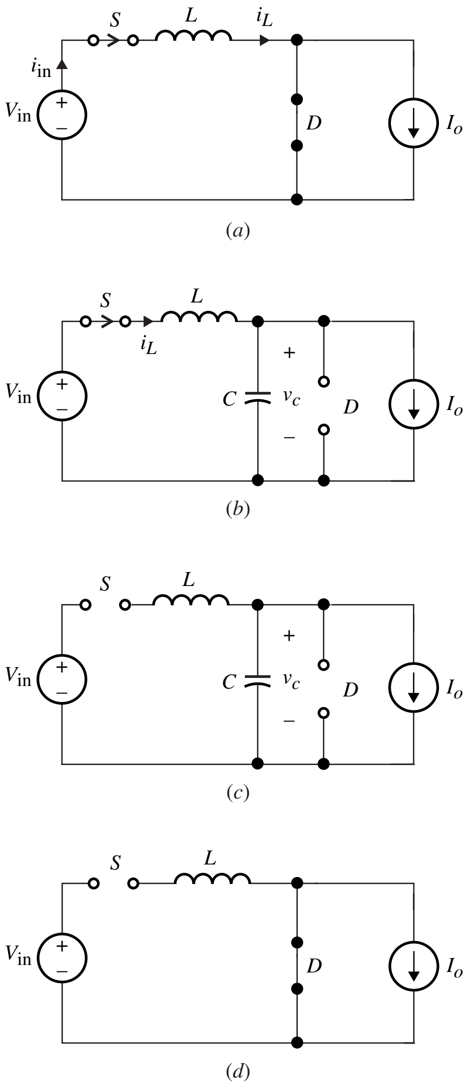

Figure 6.9 (*a*) Equivalent circuit for Mode I. (*b*) Equivalent circuit for mode II. (*c*) Equivalent circuit for mode III. (*d*) Equivalent circuit for mode IV.

As long as the inductor current is less than I_o, the diode will continue conducting and the capacitor voltage remains at zero. At time t_1, the inductor current becomes equal to I_o, the diode stops conducting, and the circuit enters mode II. Evaluating Eq. (6.2) at $t = t_1$, we have

$$I_o = \frac{V_{\text{in}}}{L} t_1 \tag{6.3}$$

Hence, the time interval $\Delta t_1 = t_1$ is given by

$$\Delta t_1 = t_1 = \frac{L I_o}{V_{\text{in}}} \tag{6.4}$$

This is the inductor current charging state.

Mode II **($t_1 \le t < t_2$)** Mode II starts at t_1, when the diode is open-circuited as shown in Fig. 6.9(*b*), resulting in a resonant stage between L and C. During the time between t_1 and t_2, the switch remains on, but the diode is off. The initial capacitor voltage remains zero, but the initial inductor current changes to I_o.

The first-order differential equations that represent this mode are

$$C\frac{dv_c}{dt} = i_L - I_o \tag{6.5a}$$

$$L\frac{di_L}{dt} = V_{\text{in}} - v_c \tag{6.5b}$$

Next we express the inductor current in a second-order differential equation for $t \ge t_1$ as given by

$$\frac{d^2 i_L}{dt^2} + \frac{1}{LC} i_L = \frac{I_o}{LC} \tag{6.6}$$

From Eq. (6.6), the general solution for $i_L(t)$ is given by

$$i_L(t) = A_1 \sin\omega_o(t - t_1) + A_2 \cos\omega_o(t - t_1) + A_3 \tag{6.7}$$

where the resonant angular frequency is defined by

$$\omega_o = \sqrt{\frac{1}{LC}}$$

Now we evaluate the constants A_1, A_2, and A_3. Recall at $t = t_1$, $i_L(t_1) = I_o$. Using this value, Eq. (6.7) becomes

$$A_2 + A_3 = I_o$$

The other relation in terms of the constants is obtained from the following equation:

$$L\frac{di_L(t)}{dt} = V_{\text{in}} - v_c(t)$$

Substituting for i_L from Eq. (6.7), and since the capacitor voltage is zero at t_1, we have

$$A_1 = \frac{V_{\text{in}}}{L\omega_o}$$

By taking the derivative of i_L again where $L(di_L^2(t_1)/dt^2) = 0$, we find A_2 to equal zero. Therefore, $A_3 = I_o$.

By solving the above equations, i_L and v_c are given by

$$i_L(t) = I_o + \frac{V_{\text{in}}}{Z_o}\sin\omega_o(t-t_1) \tag{6.8}$$

$$v_c(t) = V_{\text{in}}[1-\cos\omega_o(t-t_1)] \tag{6.9}$$

where $Z_o = \sqrt{L/C}$ is known as the characteristic impedance.

This mode will last until $t = t_2$, when the transistor turns off because the inductor current reaches zero. The peak inductor current occurs at $t = t'_{\text{max}}$, where $\omega_o(t_{\text{max}}{}' - t_1) = \pi/2$. When $v_c(t) = V_{\text{in}}$, the peak inductor current is $I_o + V_{\text{in}}/Z_o$. Moreover, the peak capacitor voltage occurs at $t = t''_{\text{max}}$, where $\omega_o(t''_{\text{max}} - t_1) = \pi$. When $i_L(t) = I_o$, the peak capacitor voltage is 2 V_{in}.

The time interval in this mode can be derived at $t = t_2$ by setting $i_L(t_2) = 0$,

$$\begin{aligned} i_L(t_2) &= I_o + \frac{V_{\text{in}}}{Z_o}\sin\omega_o(t_2-t_1) \\ &= 0 \end{aligned} \tag{6.10}$$

Therefore,

$$\Delta t_2 = t_2 - t_1 = \frac{1}{\omega_o}\sin^{-1}\frac{-Z_o I_o}{V_{\text{in}}} \tag{6.11}$$

It should be noted that the angle $\omega_o(t_2 - t_1)$ is in the third quadrant; hence, Eq. (6.11) may be written as

$$t_2 - t_1 = \frac{T_s}{2} + \frac{1}{\omega_o}\sin^{-1}\frac{Z_o I_o}{V_{\text{in}}} = \frac{1}{\omega_o}\left(\pi + \sin^{-1}\frac{Z_o I_o}{V_o}\right)$$

Mode III starts at $t = t_2$, when the switch is turned off.

Mode III **($t_2 \le t < t_3$)** At t_2 the inductor current becomes zero, and the capacitor linearly discharges from $v_c(t_2)$ to zero during t_2 to t_3. The diode remains off since its voltage is negative, as shown in Fig. 6.9(*c*).

Now the initial value of $i_L(t_2)$ is zero and the initial value of the capacitor at $t = t_2$ is $V_c(t_2)$. The inductor has no current going through it when the switch is off, so the capacitor current equals I_o, as given by

$$i_c = C\frac{dv_c}{dt} = -I_o \tag{6.12}$$

The capacitor voltage $v_c(t)$ is obtained by integrating Eq. (6.12) from t_2 to t with $V_c(t_2)$ as the initial value, resulting in

$$v_c(t) = \frac{-I_o}{C}(t-t_2) + V_c(t_2) \tag{6.13}$$

The initial value $V_c(t_2)$ is obtained from Eq. (6.9) for the previous mode, to yield

$$V_c(t_2) = V_{\text{in}}[1-\cos\omega_o(t_2-t_1)] \tag{6.14}$$

Substituting Eq. (6.14) into Eq. (6.13), we obtain

$$v_c(t) = \frac{-I_o}{C}(t-t_2) + V_{\text{in}}[1-\cos\omega(t_2-t_1)] \tag{6.15}$$

At $t = t_3$, the capacitor voltage becomes zero, and the equation for this time interval is given by

$$\Delta t_3 = t_3 - t_2 = \frac{C}{I_o} V_{\text{in}} [1 - \cos\omega_o (t_2 - t_1)] \tag{6.16}$$

At this point, the diode turns on and the circuit enters mode IV. In this mode, the capacitor voltage and inductor current remain zero until the switch is turned on again to repeat mode I.

Mode IV $(t_3 \le t < t_4)$ In this mode the switch remains off, but the diode starts conducting at $t = t_3$. Mode IV will continue as long as the switch is off, and the output current starts the free-wheeling stage through the diode. The inductor current and the capacitor voltage are zero when the switch closes.

$$i_L(t) = 0$$
$$v_c(t) = 0$$

Therefore, there will be no power transfer during this mode. By turning on the switch at $t = T_s$, the cycle will repeat these four modes. The dead time Δt_4 is given by

$$\Delta t_4 = T_s - \Delta t_1 - \Delta t_2 - \Delta t_3 \tag{6.17}$$

Figure 6.10 shows the steady-state waveforms for v_c and i_L for the buck converter with L-type switch.

Voltage Gain In this section, a detailed derivation for the expression for the voltage gain, $M = V_o / V_{\text{in}}$, in terms of the circuit parameters will be given.

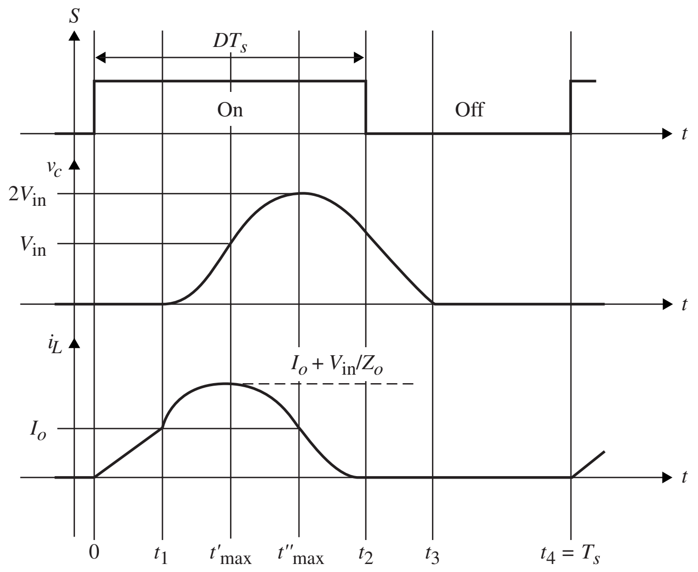

Figure 6.10 Steady-state current and voltage waveforms of buck L-type.

The average output voltage, V_o, can be obtained by evaluating the following integral.

$$V_o = \frac{1}{T_s}\int_0^{T_s} v_c(t)\,dt \tag{6.18}$$

Substitute for $v_c(t)$ from Eqs. (6.9) and (6.13) for the intervals $(t_2 - t_1)$ and $(t_3 - t_2)$, respectively, into Eq. (6.18) to yield

$$V_o = \frac{1}{T_s}\left[\int_{t_1}^{t_2} V_{\text{in}}(1 - \cos\omega_o(t - t_1))\,dt + \int_{t_2}^{t_3}\left(\frac{-I_o}{C}(t - t_2) + V_c(t_2)\right)dt\right]$$

The voltage gain ratio is given by

$$\frac{V_o}{V_{\text{in}}} = \frac{1}{T_s}\left[(t_2 - t_1) - \frac{\sin\omega_o(t_2 - t_1)}{\omega_o} - \frac{I_o}{V_{\text{in}}C}\frac{(t_3 - t_2)^2}{2} + V_c(t_2)(t_3 - t_2)\right] \tag{6.19}$$

Substitute for $(t_2 - t_1)$, $(t_3 - t_2)$, and $V_c(t_2)$ from Eqs. (6.11), (6.16), and (6.14), respectively, into Eq. (6.19) to yield a closed-form expression for M in terms of the circuit parameters. However, for illustration purposes, we will show that the same expression can be obtained from the law of power conservation. The conservation of energy per switching cycle states that since the converter is assumed to be ideal, then the average input and output powers should be equal. Next we evaluate the average input and output powers and then equate them since the converter is assumed to be ideal.

The total input energy over one switching cycle is given by

$$E_{\text{in}} = \int_0^{T_s} i_{\text{in}} V_{\text{in}}\,dt \tag{6.20}$$

Since i_{in} is equal to $i_L(t)$, Eq. (6.20) is rewritten as

$$E_{\text{in}} = \int_0^{t_1} i_L(t) V_{\text{in}}\,dt + \int_{t_1}^{t_2} i_L(t) V_{\text{in}}\,dt \tag{6.21}$$

Substituting for $i_L(t)$ from Eqs. (6.2) and (6.8) into the integrals, respectively, Eq. (6.21) becomes

$$E_{\text{in}} = V_{\text{in}}\left\{\frac{V_{\text{in}}}{2L}t_1^2 + I_o(t_2 - t_1) + \frac{V_{\text{in}}}{Z_o\omega_o}[\cos\omega_o(t_2 - t_1) - 1]\right\} \tag{6.22}$$

Substituting for $\cos\omega_o(t_2 - t_1) = 1 - [I_o(t_3 - t_2)]/CV_{\text{in}}$ from Eq. (6.16) for mode III, Eq. (6.22) becomes

$$E_{\text{in}} = V_{\text{in}}\left\{\frac{t_1}{2}I_o + I_o(t_2 - t_1) + \frac{V_{\text{in}}}{Z_o\omega_o}\left[\frac{I_o(t_3 - t_2)}{CV_{\text{in}}}\right]\right\} \tag{6.23}$$

With $Z_o\omega_o = 1/C$, Eq. (6.23) gives

$$E_{\text{in}} = V_{\text{in}}I_o\left[\frac{t_1}{2} + (t_2 - t_1) + (t_3 - t_2)\right] \tag{6.24}$$

The output energy over one cycle is obtained by evaluating Eq. (6.25):

$$E_o = \int_0^{T_s} I_o V_o\,dt = I_o V_o T_s \tag{6.25}$$

From the conservation of energy theory, equating the input and output energy expressions from Eqs. (6.24) and (6.25), we have

$$I_o V_o T_s = V_{\text{in}} I_o \left[\frac{t_1}{2} + (t_2 - t_1) + (t_3 - t_2)\right] \tag{6.26}$$

From Eq. (6.26) the voltage gain is expressed by

$$\frac{V_o}{V_{\text{in}}} = \frac{1}{T_s}\left[\frac{t_1}{2} + (t_2 - t_1) + (t_3 - t_2)\right] \tag{6.27}$$

Substituting for t_1, $(t_2 - t_1)$, and $(t_3 - t_2)$ from Eqs. (6.4), (6.11), and (6.16), respectively, into Eq. (6.27), the voltage gain becomes

$$\frac{V_o}{V_{\text{in}}} = \frac{1}{T_s}\left\{\frac{LI_o}{2V_{\text{in}}} + \frac{1}{\omega_o}\sin^{-1}\frac{-Z_o I_o}{V_{\text{in}}} + \frac{CV_{\text{in}}}{I_o}[1 - \cos\omega_o(t_2 - t_1)]\right\} \tag{6.28}$$

To simplify and generalize the gain equation, the following normalized parameters are defined:

$$M = \frac{V_o}{V_{\text{in}}} \quad \text{normalized output voltage (voltage gain)} \tag{6.29a}$$

$$Q = \frac{R_o}{Z_o} \quad \text{normalized load} \tag{6.29b}$$

$$I_o = \frac{V_o}{R_o} \quad \text{average output current} \tag{6.29c}$$

$$f_{ns} = \frac{f_s}{f_o} \quad \text{normalized switching frequency} \tag{6.29d}$$

By substituting Eqs. (6.29) into Eq. (6.28), the final voltage gain is simplified to

$$M = \frac{f_{ns}}{2\pi}\left[\frac{M}{2Q} + \alpha + \frac{Q}{M}(1 - \cos\alpha)\right] \tag{6.30}$$

where

$$\alpha = \sin^{-1}\left(\frac{-M}{Q}\right) \tag{6.31}$$

A plot of the control characteristic curve of M vs. f_{ns} under various normalized loads is given in Fig. 6.11.

EXAMPLE 6.1

Consider the following specifications for the ZCS buck converter of Fig. 6.8(*a*). Assume the parameters are $V_{\text{in}} = 25$ V, $V_o = 12$ V, $I_o = 1$ A, $f_s = 250$ kHz. Design for the resonant tank parameters L and C and calculate the peak inductor current and peak capacitor voltage. Determine the time interval for each mode.

SOLUTION The voltage gain is $M = V_o/V_{\text{in}} = 12/25 = 0.48$. Let us select $f_{ns} = 0.4$. Next we determine Q from either the control characteristic curve of Fig. 6.11 or from the gain equation of

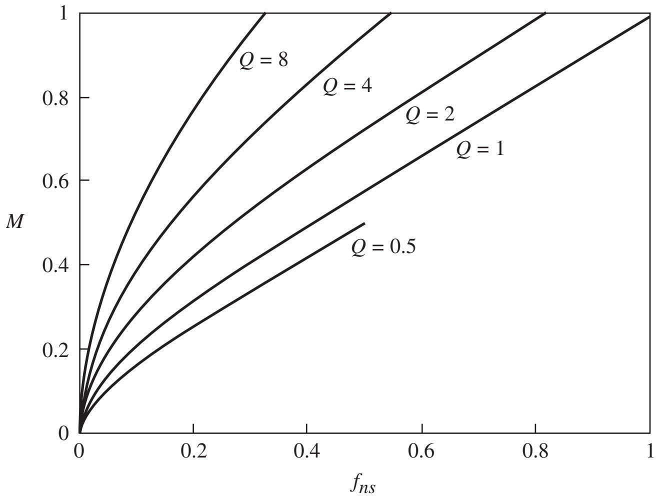

Figure 6.11 Control characteristic curve of M vs. f_{ns} for the ZCS buck converter.

Eq. (6.30). This results in Q approximately equal to 1. Since $R_o = V_o/I_o$, the characteristic impedance is given by

$$Z_o = \frac{R_o}{Q} = 12\ \Omega$$
$$= \sqrt{\frac{L}{C}} = 12\ \Omega \tag{6.32}$$

The second equation in terms of L and C is obtained from f_o. From the normalized switching frequency, f_o may be given by

$$f_o = \frac{f_s}{f_{ns}}$$
$$= \frac{f_s}{0.4} = 625\text{ kHz}$$

In terms of the angular frequency, ω_o, we have

$$\omega_o = 2\pi f_o = \sqrt{\frac{1}{LC}} \tag{6.33}$$

Solving Eqs. (6.32) and (6.33) for L and C, we obtain

$$L = \frac{Z_o}{\omega_o} = \frac{12\ \Omega}{2\pi \times 625 \times 10^3\text{ rad/s}}$$
$$= 3.06 \times 10^{-6} \approx 3\ \mu\text{H}$$
$$C = \frac{1}{Z_o \omega_o} = \frac{1}{12 \times 2 \times \pi \times 625 \times 10^3}$$
$$\approx 0.02\ \mu\text{F}$$

The peak inductor current is given by

$$I_{\text{L,peak}} = I_o + \frac{V_{\text{in}}}{Z_o}$$
$$\approx 3\text{ A}$$

and the peak capacitor voltage is

$$v_{c,\text{peak}} = 2V_{\text{in}}$$
$$= 50 \text{ V}$$

The time intervals are calculated from the following expressions:

$$t_1 = \frac{I_o L}{V_{\text{in}}} = \frac{1 \text{ A} \times (3 \times 10^{-6} \text{ H})}{25 \text{ V}} \approx 0.122 \ \mu\text{s}$$

$$t_2 = t_1 + \frac{1}{\omega_o}\sin^{-1}\left(\frac{-Z_o I_o}{V_{\text{in}}}\right)$$
$$= 0.122 + \frac{1}{2\pi \times 625 \times 10^3}\sin^{-1}\left(\frac{-12 \times 1}{25}\right)$$
$$\approx 0.795 \ \mu\text{s}$$

$$t_3 = t_2 + \frac{CV_{\text{in}}(1 - \cos\omega_o(t_2 - t_1))}{I_o}$$
$$= 0.795 + \frac{(0.02 \times 25 \times 10^{-6})(1 - \cos(2\pi \times 0.625 \times 0.67))}{1 \text{ A}}$$
$$\approx 1.79 \ \mu\text{s}$$

For $t'_{\max}$ we have

$$\omega_o(t'_{\max} - t_1) = \frac{\pi}{2}$$

$$t'_{\max} = \frac{\pi/2}{\omega_o} + t_1$$
$$= 0.4 \ \mu\text{s} + 0.122 \ \mu\text{s}$$
$$= 0.522 \ \mu\text{s}$$

and finally, $t_4 = 4 \ \mu\text{s} = T_s$.

EXERCISE 6.1

Consider the ZCS buck converter with the following parameters: $V_{\text{in}} = 40$ V, $V_o = 28.7$ V @ $I_o = 0.6$ A, $f_s = 100$ kHz, $L = 15$ μH, and $C = 60$ nF. Determine the peak inductor current and the time at which the capacitor reaches its maximum voltage of 40 V.

ANSWER 3.1 A, 3.2 μs

EXERCISE 6.2

Figure 6.12(*a*) shows another type of quasi-resonant buck converter that uses an M-type resonant switch arrangement. Its simplified circuit is shown in Fig. 6.12(*b*). Derive the voltage gain equation, $M = V_o/V_{\text{in}}$, for the steady-state waveforms shown in Fig. 6.12(*c*).

The ZCS Boost Converter

Let us consider the boost quasi-resonant converter with an M-type switch as shown in Fig. 6.13(*a*), with its equivalent circuit shown in Fig. 6.13(*b*). Here we assume the input

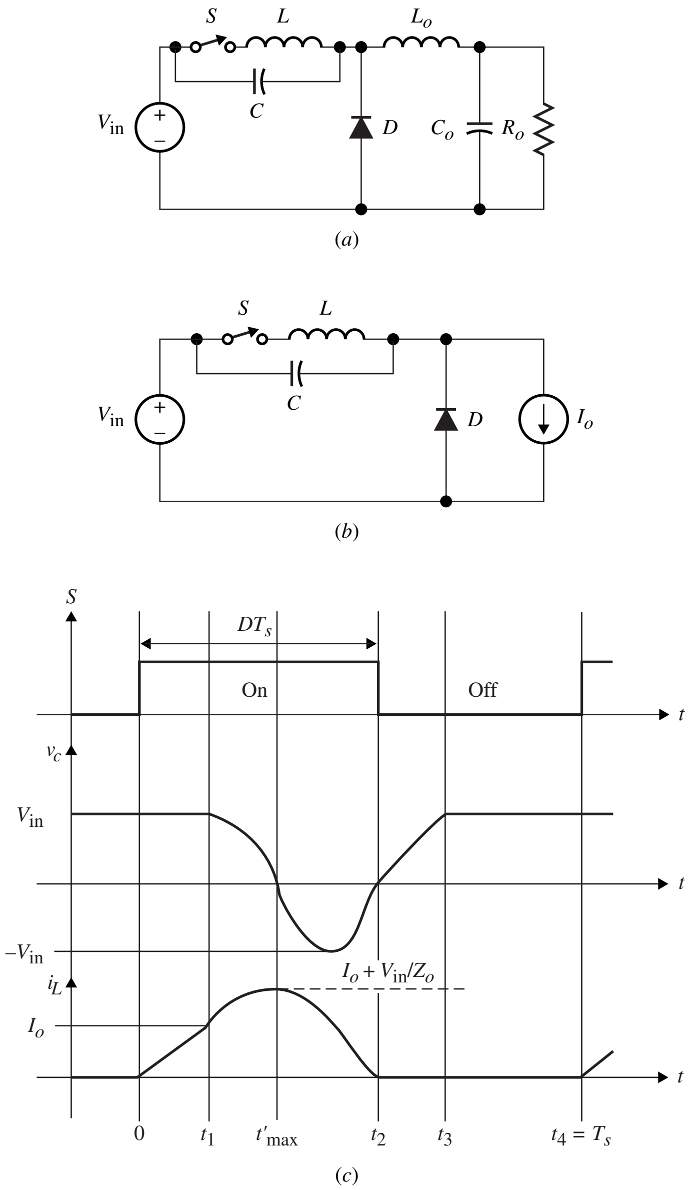

Figure 6.12 (*a*) ZCS buck converter with M-type switch. (*b*) Simplified equivalent circuit. (*c*) Steady-state waveforms.

current is constant and the load voltage is constant. The following analysis assumes half-wave operation.

Mode I **($0 \le t < t_1$)** We first assume that the switch and the diode are both on in mode I, as shown in Fig. 6.14(*a*). The output voltage is given by

$$V_o = L\frac{di_L}{dt} \tag{6.34}$$

The initial inductor current and capacitor voltage are given by

$$i_L(0) = 0$$
$$v_c(0) = V_o$$

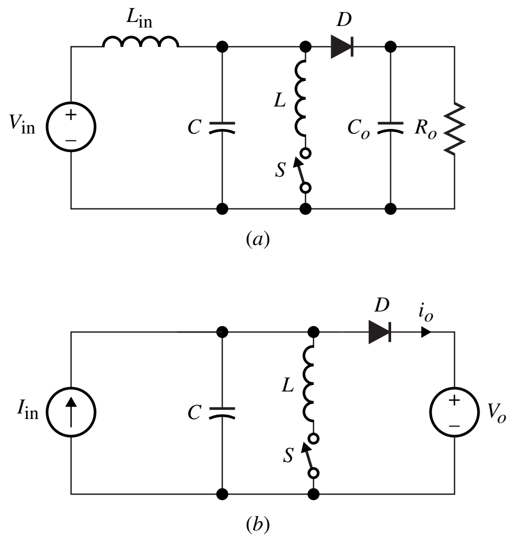

Figure 6.13 (*a*) ZCS boost converter with M-type switch. (*b*) Simplified equivalent circuit.

Integrating Eq. (6.34), the inductor current becomes

$$i_L(t) = \frac{V_o}{L}t + i_L(0) = \frac{V_o}{L}t \tag{6.35}$$

When the resonant inductor current reaches the input current, I_{in}, the diode turns off; hence, we have

$$\frac{V_o}{L}t_1 = I_{in}$$

with t_1 given by

$$t_1 = \frac{I_{in}L}{V_o} \tag{6.36}$$

At $t = t_1$, the diode turns off since $i_L = I_{in}$, and the converter enters mode II.

Mode II ($t_1 \le t < t_2$) The switch remains closed, but the diode is off at t_1 in mode II as shown in Fig. 6.14(*b*). This is a resonant mode during which the capacitor voltage starts decreasing resonantly from its initial value of V_o. When $i_L = I_{in}$, the capacitor reaches its negative peak. At $t = t_2$, i_L equals zero, and the switch turns off (i.e., switching at zero current).

The initial conditions are given by

$$v_c(t_1) = V_o \quad \text{and} \quad i_L(t_1) = I_o$$

From Fig. 6.14(*b*), the first derivatives for i_L and v_c are given by

$$L\frac{di_L}{dt} = v_c(t)$$

$$C\frac{dv_c}{dt} = I_{in} - i_L(t)$$

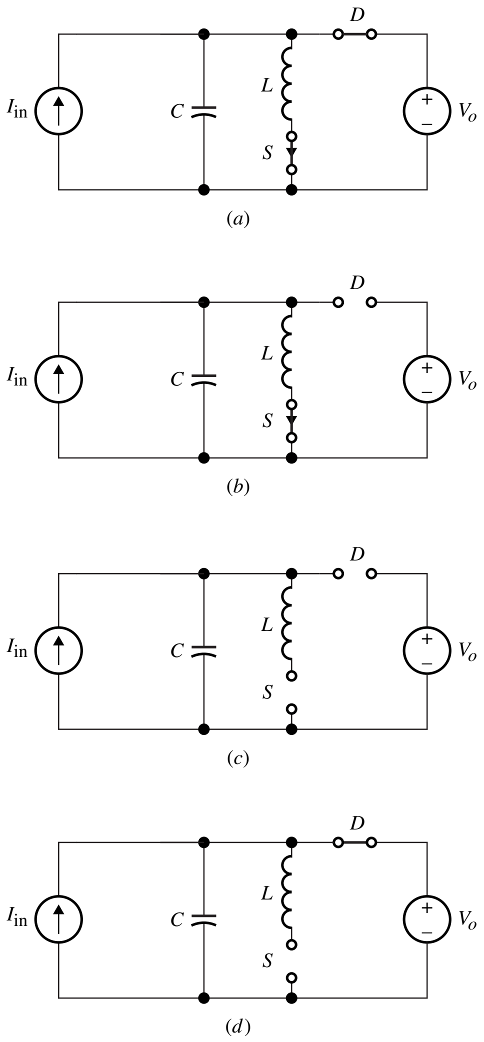

Figure 6.14 (*a*) Equivalent circuit for mode I. (*b*) Equivalent circuit for mode II. (*c*) Equivalent circuit for mode III. (*d*) Equivalent circuit for mode IV.

Next the two first-order differential equations need to be solved for this mode. Using the same solution technique used with the buck converter, the expressions for $i_L(t)$ and $v_c(t)$ are given by

$$i_L(t) = I_{\text{in}} + \frac{V_o}{Z_o}\sin\omega_o(t - t_1) \tag{6.37}$$

$$v_c(t) = V_o\cos\omega_o(t - t_1) \tag{6.38}$$

where $\omega_o = 1/\sqrt{LC}$.

At $t = t_2$, $i_L(t_2) = 0$, and the time interval can be obtained from evaluating Eq. (6.37) at $t = t_2$, to yield

$$\begin{aligned}(t_2 - t_1) &= \frac{1}{\omega_o}\sin^{-1}\left(-\frac{I_{\text{in}}Z_o}{V_o}\right) \\ &= \frac{1}{\omega_o}\left[\pi + \sin^{-1}\left(\frac{I_{\text{in}}Z_o}{V_o}\right)\right]\end{aligned} \tag{6.39}$$

Mode III **($t_2 \le t < t_3$)** Mode III starts at t_2, and the switch and the diode are both open, as shown in Fig. 6.14(*c*). Since v_c is constant, the capacitor starts charging up by the input current source. The capacitor voltage is given by

$$\begin{aligned} v_c(t) &= \frac{1}{C}\int_{t_2}^{t} I_{\text{in}}\,dt \\ &= \frac{I_{\text{in}}}{C}(t - t_2) + v_c(t_2) \end{aligned} \tag{6.40}$$

The diode begins conducting at $t = t_3$ when the capacitor voltage is equal to the output voltage, i.e., $v_c(t_3) = V_o$. Equation (6.40) becomes

$$v_c(t_3) = V_o = \frac{I_{\text{in}}}{C}(t_3 - t_2) + v_c(t_2)$$

so the time interval in this period can be expressed as

$$t_3 - t_2 = \frac{I_{\text{in}}}{C}[V_o - v_c(t_2)] \tag{6.41}$$

where $v_c(t_2)$ may be obtained from Eq. (6.38).

Mode IV **($t_3 \le t < t_4$)** At t_3 the capacitor voltage is clamped to the output voltage and the diode starts conducting again, but the switch remains open. This condition remains as long as the switch is open. The cycle of the mode will repeat at the time of T_s, when S is turned on again.

Typical steady-state waveforms are shown in Fig. 6.15.

Voltage Gain As we did for the buck converter, we apply conservation of energy per switching cycle to express the voltage gain, $M = V_o/V_{\text{in}}$, in terms of the circuit parameter.

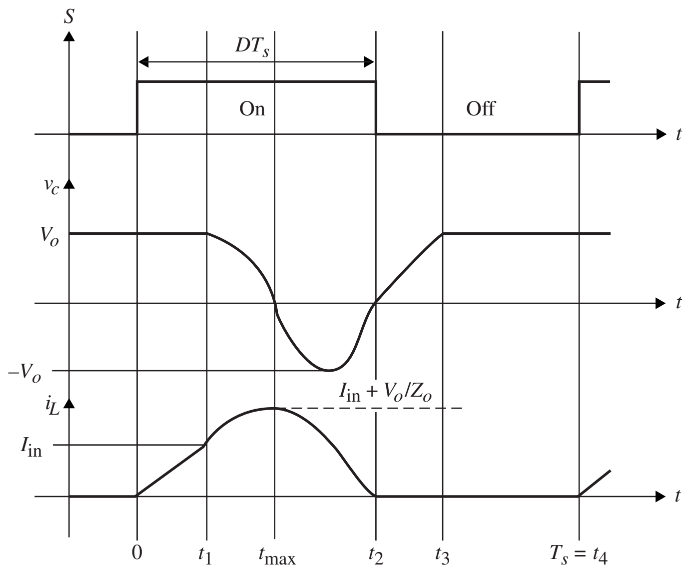

Figure 6.15 Steady-state waveforms of the boost converter with M-type switch.

The input energy is given by

$$E_{\text{in}} = V_{\text{in}} I_{\text{in}} T_s \tag{6.42}$$

and the output energy by

$$E_o = \int_0^{T_s} i_o V_o \, dt \tag{6.43}$$

The output current equals $i_o = I_{\text{in}} - i_L$ and $i_o = I_{\text{in}}$ for intervals $0 \le t < t_1$ and $t_3 \le t < T_s$, respectively. Therefore, E_o becomes

$$E_o = \int_0^{t_1} (I_{\text{in}} - i_L) V_o \, dt + \int_{t_3}^{T_s} I_{\text{in}} V_o \, dt \tag{6.44}$$

The input current is obtained from the conservation of output power as

$$I_{\text{in}} = \frac{V_o^2}{V_{\text{in}} R_o}$$

Substituting for the input current and evaluating Eq. (6.44), the output energy becomes

$$\begin{aligned} E_o &= V_o \int_0^{t_1} \left(I_{\text{in}} - \frac{V_o}{L} t \right) dt + I_{\text{in}} V_o (T_s - t_3) \\ &= V_o \left(I_{\text{in}} t_1 - \frac{1}{2} \frac{V_o}{L} t_1^2 \right) + I_{\text{in}} V_o (T_s - t_3) \end{aligned} \tag{6.45}$$

If we substitute for $t_1 = I_{\text{in}} L / V_o$ and $(T_s - t_3) = T_s - [t_1 + (t_2 - t_1) + (t_3 - t_2)]$, and use the equations for $(t_2 - t_1)$ and $(t_3 - t_2)$ from Eqs. (6.39) and (6.41), Eq. (6.45) becomes

$$E_o = -\frac{1}{2} I_{\text{in}}^2 L + V_o I_{\text{in}} \left[T_s - \frac{\alpha}{\omega_o} - \frac{C}{I_{\text{in}}} V_o (1 - \cos\alpha) \right] \tag{6.46}$$

Following similar steps as for the quasi-resonant buck converter, it can be shown that the voltage gain expression is given by

$$\frac{M-1}{M} = \frac{f_{ns}}{2\pi} \left[\frac{M}{2Q} + \alpha + \frac{Q}{M} (1 - \cos\alpha) \right] \tag{6.47}$$

where, α, M, I_o, and f_{ns} are given in Eqs. (6.29).

Using Eq. (6.47), Fig. 6.16 shows the characteristic curve for M vs. f_{ns} as a function of the normalized load.

EXAMPLE 6.2

Design a boost ZCS converter for the following parameters: $V_{\text{in}} = 20$ V, $V_o = 40$ V, $P_o = 20$ W, $f_s = 250$ kHz.

SOLUTION The voltage gain is $M = V_o / V_{\text{in}} = 40/20 = 2$. Let us select $f_{ns} = 0.38$. From the characteristic curve of Fig. 6.16, Q can be approximated as 6.0.

The characteristic impedance is given by

$$Z_o = \frac{R_o}{Q} = \frac{V_o^2 / P_o}{Q} = \frac{80\ \Omega}{6} = 13.33\ \Omega \tag{6.48}$$

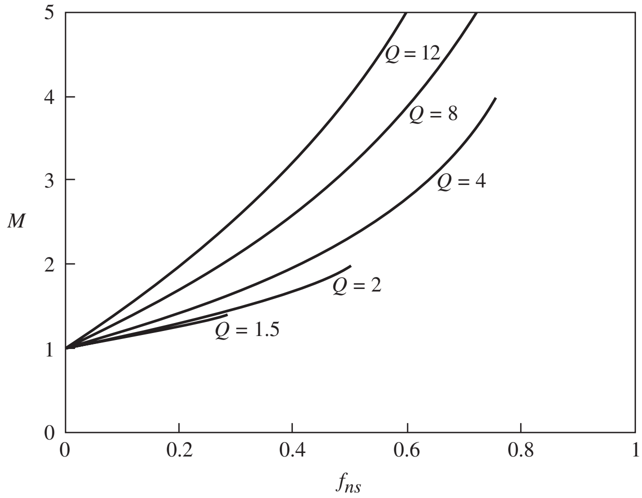

Figure 6.16 Characteristic curve for M vs. f_{ns} for the boost ZCS converter.

and the resonant frequency is

$$\begin{aligned} f_o &= \frac{f_s}{f_{ns}} \\ &= \frac{250 \text{ kHz}}{0.38} = 657.89 \text{ kHz} \end{aligned} \tag{6.49}$$

Solve Eqs. (6.48) and (6.49) for L and C:

$$L = \frac{Z_o}{2\pi f_o} = \frac{13.33\ \Omega}{2\pi \times 657.89 \times 10^3} = 3.22 \times 10^{-6} \text{ H}$$

$$C = \frac{1}{Z_o \omega_o} = \frac{1}{(13.33)(2\pi \times 657.89 \times 10^3)} = 18.14 \text{ nF}$$

To limit the input ripple current and the output voltage, we set

$$L_o = 100L = 322 \times 10^{-6} \text{ H}$$

$$C_o = 100C = 1.8 \times 10^{-6} \text{ F}$$

EXAMPLE 6.3

Design a boost converter with ZCS, with the following design parameters: $V_{\text{in}} = 25$ V, $P_o = 30$ W at $I_o = 0.5$ A, and $f_s = 100$ kHz. Assume the output voltage ripple $\Delta V_o / V_o$ is 0.2%.

SOLUTION The load resistance, $R_o = P_o / I_o^2 = 30/(0.5)^2 = 120\ \Omega$.

$$M = \frac{V_o}{V_{\text{in}}} = \frac{60}{25} = 2.4$$

From the characteristic curve of Fig. 6.15, we approximate Q at 6 when we assume $f_n = 0.58$. Hence, $f_o = 100 \text{ kHz}/0.58 = 172.4$ kHz.

The characteristic impedance is obtained from

$$Q = \frac{R_o}{Z_o} = \frac{120}{Z_o} = 6 \quad \text{and} \quad Z_o = 20\ \Omega$$

Hence, $\sqrt{L/C} = 20\ \Omega$.

$$\omega_o = 2\pi(172 \times 10^3) = 1080.7 \times 10^3\ \text{rad/s}$$

$$\sqrt{\frac{1}{LC}} = 1080.7 \times 10^3\ \text{rad/s}$$

Solving for C and L,

$$C = 46.27\ \text{nF}$$

$$L = 18.51\ \mu\text{H}$$

From $\Delta V_o/V_o = 0.2\%$, C_o can be obtained from the ripple voltage equation for the conventional boost converter, which is given by

$$\frac{\Delta V_o}{V_o} = \frac{D}{f_s R_o C_o}$$

Solving the above equation for C_o we obtain

$$C_o = \frac{D}{f_s R_o (\Delta V_o/V_o)}$$

$$= \frac{((10 - 6.147 - 0.193)/10) \times 10^{-6}}{(100 \times 10^3)(120)(0.2/100)} = 15.25\ \mu\text{F}$$

The time intervals are given by

$$t_1 = \frac{L I_{\text{in}}}{V_o} = \frac{18.51 \times 10^{-3} \times 1.2}{60} = 0.370\ \mu\text{s}$$

$$t_2 - t_1 = \frac{1}{\omega_o}\sin^{-1}\left[\frac{-Z_o I_{\text{in}}}{V_o}\right]$$

$$= \frac{1}{1080.7 \times 10^3}\sin^{-1}\left[-\frac{20 \times 1.2}{60}\right] = 3.29\ \mu\text{s}$$

$$t_3 - t_2 = \frac{1}{\omega_n}\frac{V_o}{Z_o I_{\text{in}}}(1 - \cos\alpha)$$

$$= \frac{1}{1080.7 \times 10^3}\ \frac{60}{20 \times 1.2}(1 - \cos 3.553)$$

$$= 0.193\ \mu\text{s}$$

$$t_4 - t_3 = T - t_1 - (t_2 - t_1) - (t_3 - t_2)$$

$$= 10 - 0.370 - 3.29 - 0.193 = 6.147\ \mu\text{s}$$

The output inductor is obtained from

$$L_{\text{crit}} = \frac{R_o}{2 f_s}(1 - D^2)D$$

$$= 88.8\ \mu\text{H}$$

Hence, we select $L_o = 890\ \mu\text{H}$.

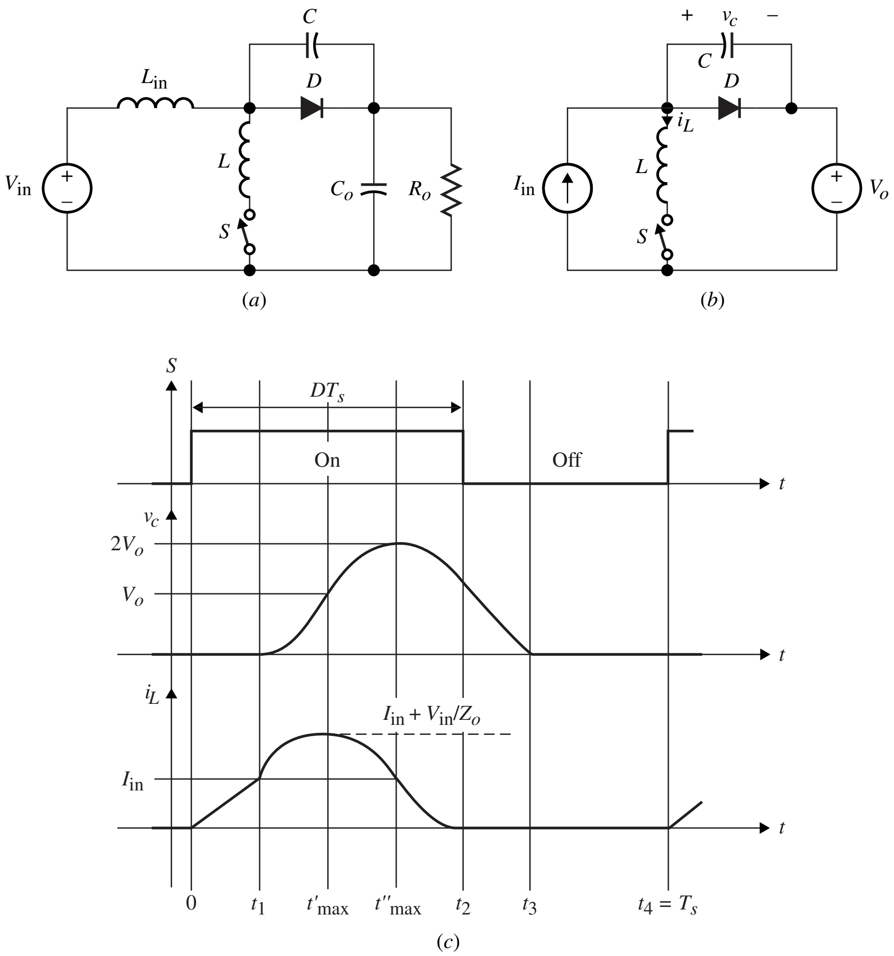

Figure 6.17 (*a*) ZCS boost converter with L-type switch. (*b*) Simplified equivalent circuit. (*c*) Steady-state waveforms.

Figure 6.17(*a*) shows the quasi-resonant boost converter using the L-type resonant switch, and the simplified circuit and steady-state waveforms are shown in Fig. 6.17(*b*) and (*c*), respectively. The reader is invited to verify these waveforms.

EXERCISE 6.3

Consider the quasi-resonant boost converter of Fig. 6.13(*a*) with the following design parameters: $V_{in} = 25$ V, $P_o = 30$ W at $I_o = 0.5$ A, and $f_s = 100$ kHz. Determine the resonant tank capacitor and inductor components, assuming $f_{ns} = 0.4$.

ANSWER $C = 42.4\ \mu$F, $L = 9.55\ \mu$H

ZCS Buck-Boost Converter

Let us consider the quasi-resonant buck-boost converter using the L-type switch as shown in Fig. 6.18(*a*); Fig. 6.18(*b*) shows the simplified equivalent circuit.

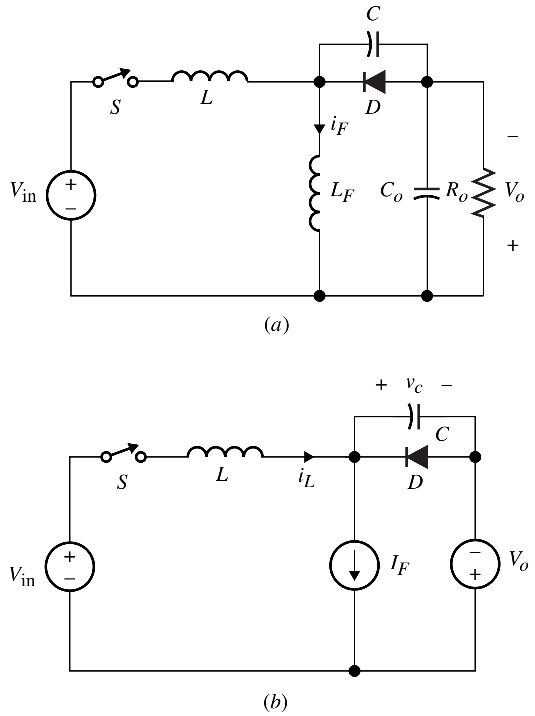

Figure 6.18 (*a*) ZCS buck-boost converter with L-type switch. (*b*) Simplified equivalent circuit.

Like the previous buck and the boost converters, the buck-boost converter has four modes of operation.

Mode I $(\mathbf{0 \le t < t_1})$ Mode I starts at $t = 0$, when the switch and the diode are both conducting. According to Kirchhoff's law, the voltage equation can be written as

$$L\frac{di_L(t)}{dt} = V_{\text{in}} + V_o \tag{6.50}$$

By integrating both sides of Eq. (6.50) with the initial condition of $i_L(0) = 0$, $i_L(t)$ is given by

$$i_L(t) = \frac{V_{\text{in}} + V_o}{L}t \tag{6.51}$$

and $v_c(t) = 0$.

At $t = t_1$, the inductor current reaches I_F, forcing the output diode to stop conducting, so t_1 can be expressed as

$$t_1 = \frac{LI_F}{V_{\text{in}} + V_o} \tag{6.52}$$

Mode II **($t_1 \le t < t_2$)** This is a resonant stage between L and C with the initial conditions given by

$$v_c(t_1) = 0$$
$$i_L(t_1) = I_F$$

Applying Kirchhoff's law in Fig. 6.19(*b*), the inductor current and capacitor voltage equations may be given as

$$\frac{di_L}{dt} = V_{\text{in}} + V_o - v_c(t) \tag{6.53a}$$

$$C\frac{dv_c}{dt} = i_L(t) - I_F \tag{6.53b}$$

Solving Eqs. (6.53) for $t > t_1$, we obtain

$$i_L(t) = I_F + \frac{V_{\text{in}} + V_o}{Z_o}\sin\omega_o(t - t_1) \tag{6.54}$$

$$v_c(t) = (V_{\text{in}} + V_o)[1 - \cos\omega_o(t - t_1)] \tag{6.55}$$

At $t = t_2$, the inductor current reaches zero, $i_L(t_2) = 0$, and the switch stops conducting. The time interval $(t_2 - t_1)$ is given by

$$(t_2 - t_1) = \frac{1}{\omega_o}\sin^{-1}\left(-\frac{I_F Z_o}{V_{\text{in}} + V_o}\right) \tag{6.56}$$

Mode III **($t_2 \le t < t_3$)** Mode III starts at $t = t_2$, when the inductor current reaches zero. The switch and the diode are both off. The capacitor starts to discharge until it reaches zero, and the diode will start to conduct again at $t = t_3$. During this period, the inductor current is zero.

$$v_c(t) = \frac{-1}{C}\int_{t_2}^{t} I_F\,dt = \frac{-I_F}{C}(t - t_2) + V_c(t_2) \tag{6.57}$$

The diode begins to conduct at the end of this mode, $t = t_3$, because the capacitor voltage is equal to zero, yielding

$$0 = \frac{-I_F}{C}(t_3 - t_2) + V_c(t_2)$$

where $V(t_2)$ may be obtained from Eq. (6.55) by evaluating it at $t = t_2$. The expression from Eq. (6.57) for the time between t_2 and t_3 is

$$(t_3 - t_2) = \frac{C}{I_F}V_c(t_2) \tag{6.58}$$

Mode IV **($t_3 \le t < t_4$)** Between t_3 and t_4, the switch remains off, but the diode is on. At the end of the cycle, the switch is closed again when the current becomes zero. The cycle of the modes will repeat at T_s.

The steady-state waveforms shown in Fig. 6.20 are the characteristic waveforms for the switch, v_c, and i_L.

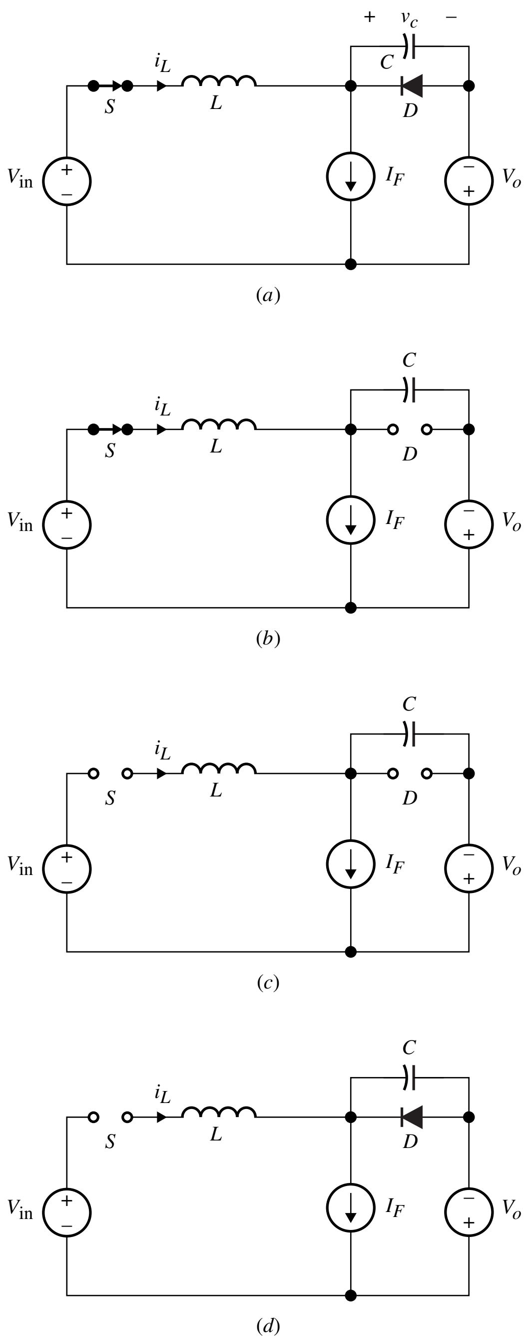

Figure 6.19 (*a*) Equivalent circuit for mode I. (*b*) Equivalent circuit for mode II. (*c*) Equivalent circuit for mode III. (*d*) Equivalent circuit for mode IV.

Voltage Gain As before, the conservation of energy per switching cycle is used to express the voltage gain, $M = V_o/V_{in}$, in terms of the normalized circuit parameter. It can be shown that M for the buck-boost ZCS converter is given by

$$\frac{M}{1+M} = \frac{f_{ns}}{2\pi}\left[\frac{M}{2Q} + \alpha + \frac{Q}{M}(1 - \cos\alpha)\right] \tag{6.59}$$

All the parameters of Eq. (6.59) are the same as before.

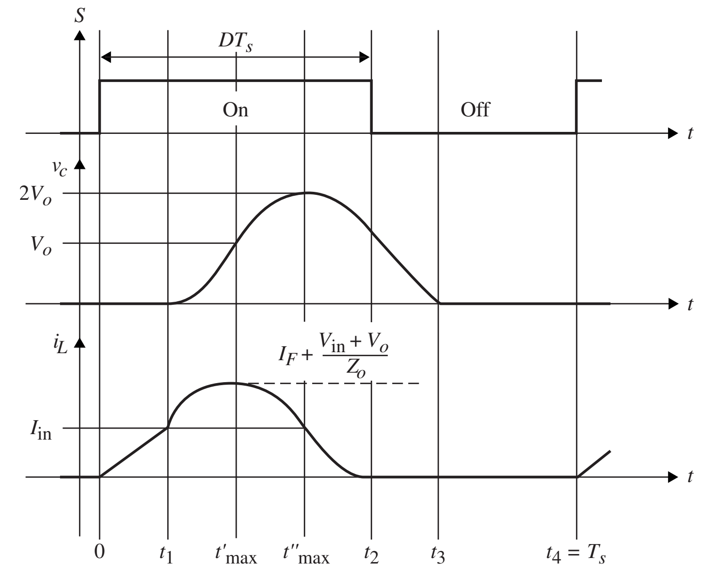

Figure 6.20 Steady-state waveforms for buck-boost converter with L-type switch.

Figure 6.21 shows the characteristic curve for M vs. f_{ns} for the ZCS buck-boost converter.

EXAMPLE 6.4

Consider a buck-boost quasi-resonant ZCS converter with the following specifications: $V_{in} = 40$ V, $P_o = 80$ W at $I_o = 4$ A, $f_s = 250$ kHz, $L_o = 0.1$ mH, and $C_o = 6\ \mu$F. Design values for L and C, and determine the output ripple voltage, assuming $D = 0.5$.

SOLUTION The output voltage and load resistance are given by

$$V_o = \frac{80}{4} = 20 \text{ V} \quad \text{and} \quad R_o = \frac{20}{4} = 5\ \Omega$$

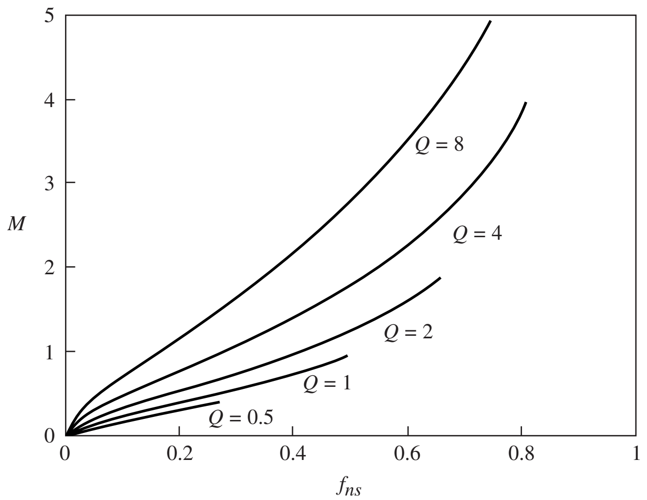

Figure 6.21 Characteristic curve for M vs. f_{ns} for the ZCS buck-boost converter.

The voltage gain is given by

$$\frac{V_o}{V_{\text{in}}} = \frac{20}{40} = 0.5$$

With $M = 0.5$ and $f_{ns} = 0.17$, we have $Q = 3$, resulting in $f_o = 250\text{ kHz}/0.17 = 1470.6\text{ kHz}$.

From Q, and Z_o, we have

$$Q = \frac{R_o}{Z_o} = \frac{5}{Z_o} \quad \text{and} \quad Z_o = \frac{3}{5} = 0.6\ \Omega$$

Hence,

$$\sqrt{L/C} = 0.6\ \Omega$$

and

$$\sqrt{1/LC} = 2\pi f_o = 2\pi \times 1470.6 \times 10^3\ \text{rad/s}$$

$$\frac{1}{C} = 0.6 \times 2\pi \times 1470.6 \times 10^3$$

From these equations C and L are given by

$$C = 180.4\ \text{nF}$$

$$L = 3^2 \times C = 1.6\ \mu\text{H}$$

The duty cycle D is approximately 33% since the voltage gain for the buck-boost is 0.5. Hence, the voltage ripple is

$$\frac{\Delta V_o}{V_o} = \frac{D}{R_o C_o f} = \frac{0.33}{5 \times 6 \times 10^{-6} \times 250 \times 10^3} = 4.4\%$$

EXERCISE 6.4

The buck-boost converter with an M-type switch is shown in Fig. 6.22(*a*), and Fig. 6.22(*b*) is the equivalent circuit. Derive the steady-state waveforms and show that the voltage gain, $M = V_o/V_{\text{in}}$, is given by

$$\frac{M}{1+M} = \frac{f_{ns}}{2\pi}\left[\frac{M}{2Q} + \alpha + \frac{Q}{M}(1 - \cos\alpha)\right] \tag{6.60}$$

6.5 ZERO-VOLTAGE SWITCHING TOPOLOGIES

In this section, we will investigate the zero-voltage switching (ZVS) quasi-resonant converter family. Like the ZCS topologies, M-type or L-type switch arrangements can be used. In those topologies, the power switch is turned on at zero voltage (of course, turn-off also occurs at zero voltage). While the switch is off, a peak voltage will appear across it, causing the stress to be higher than in the hard-switching PWM case. In a ZVS topology, a flyback diode across the switch (body diode) is used to damp the voltage across the capacitor, which results in a zero voltage across the switch. We should point out that the capacitor across the switch can be the same as the switch's parasitic capacitor, and the flyback diode could be the same as the internal body diode of the power semiconductor switch.

Figure 6.23(*a*) shows a MOSFET switch implementation including an internal body diode and a parasitic capacitor.

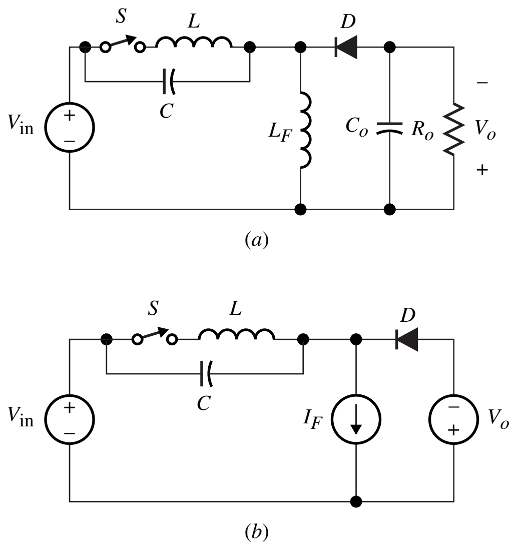

Figure 6.22 (*a*) ZCS buck-boost converter with an M-type switch. (*b*) Simplified equivalent circuit.

We will assume C_{gd} and C_{gs} are too small to be included. If the body diode, D_s, is not fast enough for the designed application or has limited power capabilities, it is possible to block it and use an external, fast flyback diode as shown in Fig. 6.23(*b*). D_1 is used to block D_s, and D_2 is the diode actually used to carry the reverse switch current. Both the current- and voltage-mode control methods are used in conjunction with a direct duty ratio PWM control approach to vary the on or off time of the power switch.

Over the last 15 years, many different zero-voltage resonant converter topologies have been introduced. Only the quasi-resonant soft-switching ZVS topologies will be analyzed and their control characteristic curves studied here. Regardless of their topological variations, many of these converters have several common features.

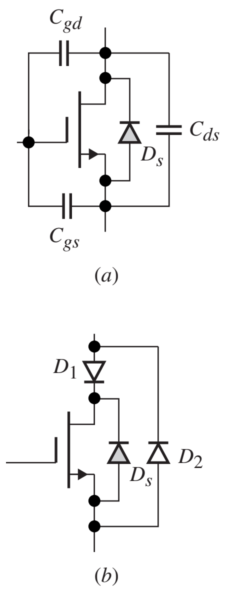

Figure 6.23 (*a*) MOSFET implementation. (*b*) MOSFET switch with fast flyback diode.

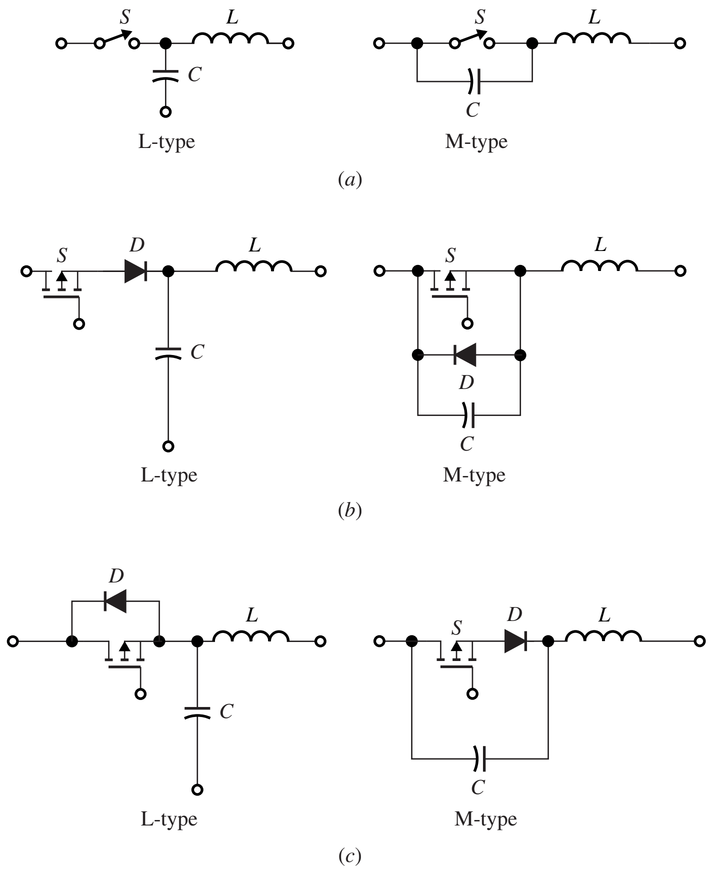

Figure 6.24 (*a*) Resonant switch arrangement types for ZVS operation. (*b*) Half-wave MOSFET implementation. (*c*) Full-wave MOSFET implementation.

6.5.1 Resonant Switch Arrangements

Next we investigate the buck, boost, and buck-boost ZVS topologies using L-type and M-type resonant switches. Figure 6.24(*a*) shows the two possible implementations using L- and M-type resonant switches. The half-wave L-type and M-type MOSFET implementations are shown in Fig 6.24(*b*), and Fig 6.24(*c*) shows the full-wave implementations for L- and M-type switches.

6.5.2 Steady-State Analyses of Quasi-Resonant Converters

As in the ZCS case, to simplify the steady-state analysis, we make the same assumptions made in Section 6.4.2.

The Buck Converter

Replacing the switch in Fig. 6.7(*a*) with the M-type switch of Fig. 6.24(*a*), we obtain the ZVS buck converter shown in Fig. 6.25(*a*). The simplified equivalent circuit is given in Fig. 6.25(*b*).

As the switch and diode are both on or off at the same time, it can be shown that under steady-state conditions there are four modes of operation. Unlike ZCS topolo-

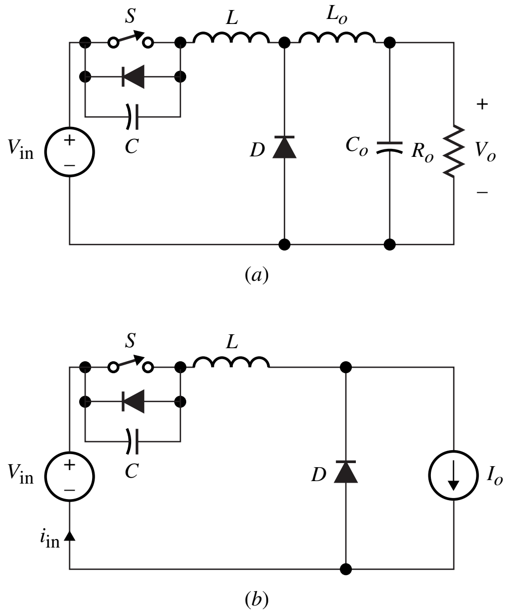

Figure 6.25 (*a*) Quasi-resonant buck converter with M-type switch. (*b*) Simplified equivalent circuit.

gies, the switching cycle in ZVS starts with the main switch in the nonconduction state. This is in order to establish a zero-voltage condition across the switch during the resonant stage while it is open. Figure 6.26 shows the equivalent circuit modes under the steady-state condition.

Mode I $(\mathbf{0 \le t < t_1})$ Assume that initially the power switch is conducting and the diode is off. Mode I starts at $t = 0$, when the switch is turned off. In this mode, since S has been closed for $t < 0$, the initial capacitor voltage, v_c, is zero, and the inductor current is I_o as shown in Fig. 6.26(*a*).

$$v_c(0) = 0$$

$$i_L(0) = I_o$$

Applying KCL to Fig. 6.26(*a*), we have

$$C\frac{dv_c}{dt} = i_L(t)$$

Since $i_L(t) = I_o$, the capacitor starts to charge according to

$$v_c(t) = \frac{1}{C}I_o t \qquad (6.61)$$

The voltage across the output diode is given by

$$v_D(t) = V_{in} - v_c(t)$$

As long as $v_c < V_{in}$, the diode remains off.

The capacitor voltage reaches the input voltage, V_{in}, at $t = t_1$, causing the diode to turn on. Hence, at $t = t_1$, we have

$$V_c(t_1) = V_{in}$$

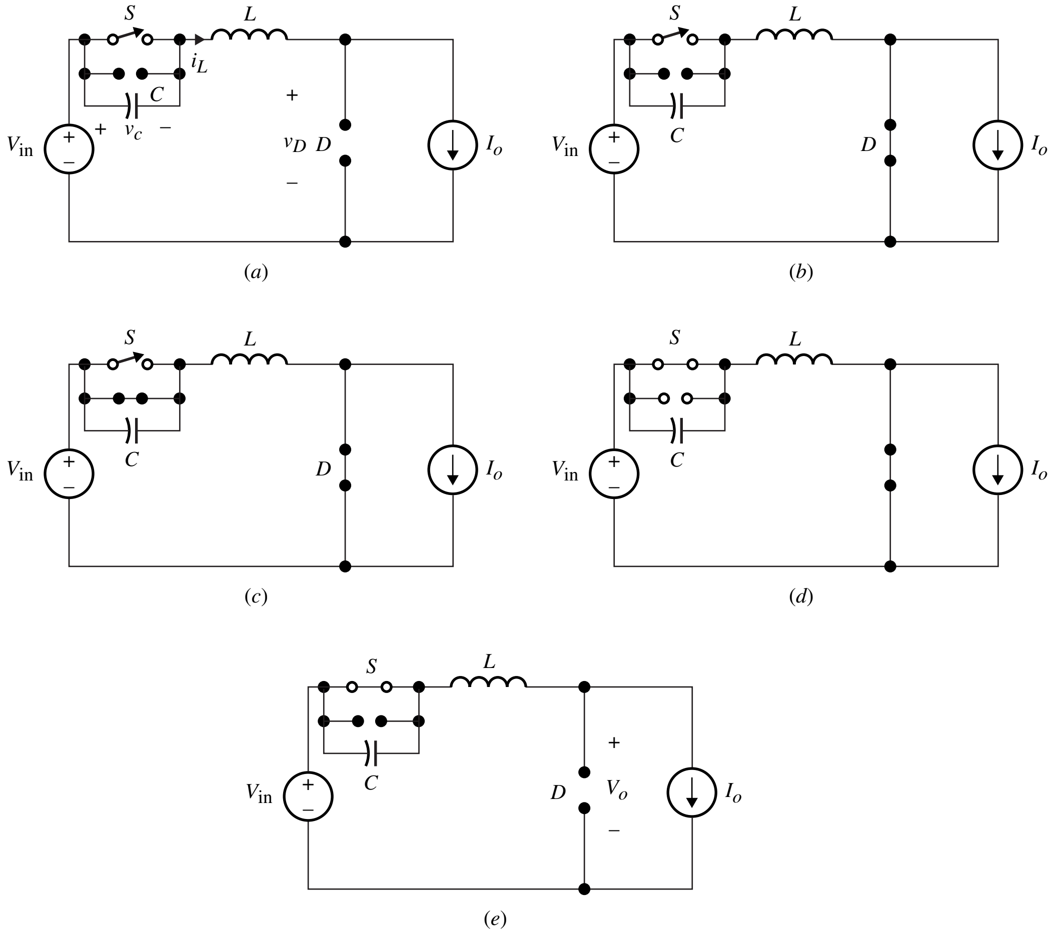

Figure 6.26 Equivalent circuits for (*a*) mode I, (*b*) mode II, (*c*) mode III $(t_2 \le t < t_2')$, (*d*) mode III $(t_2' \le t < t_3)$, and (*e*) mode IV.

and t_1 can be expressed as

$$t_1 = \frac{CV_{in}}{I_o} \tag{6.62}$$

At $t = t_1$, the circuit enters mode II. The current and voltage waveforms are shown in Fig. 6.27.

Mode II $(t_1 \le t < t_2)$ Mode II starts at t_1, when the diode turns on, and the circuit enters the resonant stage. During the time between t_1 and t_2, the switch remains off. At $t = t_2$, the capacitor voltage tends to go negative, forcing the diode across *S* to turn on. The initial capacitor voltage and inductor current in this mode are given by

$$V_c(t_1) = V_{in}$$

$$I_L(t_1) = I_o$$

The expressions for the current and the voltage in the time domain are given in Eqs. (6.63) and (6.64), respectively:

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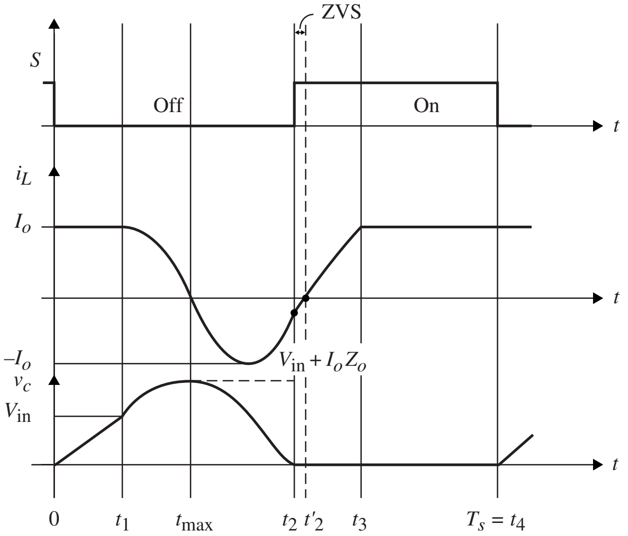


Figure 6.27 Steady-state waveforms for the ZVS buck converter.

$$i_L(t) = I_o \cos \omega_o(t - t_1) \tag{6.63}$$

$$v_c(t) = V_{in} + I_o Z_o \sin \omega_o(t - t_1) \tag{6.64}$$

where the resonant frequency and the characteristic impedance are defined as before.

The inductor current is zero when the capacitor voltage reaches the peak, and the capacitor starts discharging while the inductor current is a negative value. The inductor current reaches the peak when the capacitor drops to the input voltage, and at the end of the mode (i.e., at $t = t_2$), the capacitor voltage equals zero, $v_c(t_2) = 0$.

The period between t_2 and t_1 is given by

$$\begin{aligned} t_2 - t_1 &= \frac{1}{\omega_o}\sin^{-1}\left(\frac{-V_{in}}{I_o Z_o}\right) \\ &= \frac{\alpha}{\omega_o} \end{aligned} \tag{6.65}$$

and the inductor current at $t = t_2$ is

$$I_L(t_2) = I_o \cos \alpha \tag{6.66}$$

where

$$\alpha = \omega_o(t_2 - t_1)$$

Mode III **($t_2 \le t < t_3$)** At t_2 the capacitor voltage becomes zero, and the inductor current starts to charge linearly and reaches the output current at $t = t_3$. The body diode of the switch turns on at $t = t_2$ to maintain inductor current continuity, and the output diode also remains on at this point, as shown in Fig. 6.26(c). As long as the inductor current is less than I_o, the output diode will stay on. The switch may be turned on at 2 V any time after t_2 and before t_2', when the inductor current reverses polarity.

The initial value of the capacitor voltage in mode III is zero:

$$V_c(t_2) = 0$$

The inductor voltage is equal to the input voltage,

$$L\frac{di_L}{dt} = V_{\text{in}} \tag{6.67}$$

By integrating Eq. (6.67) from t_2 to t, the inductor current can be expressed as

$$i_L(t) = \frac{V_{\text{in}}}{L}(t - t_2) + I_o \cos\alpha \tag{6.68}$$

At $t = t_3$, the inductor current reaches the output current, $i_L(t_3) = I_o$, forcing the diode to turn off. The time interval from t_3 to t_2 is

$$t_3 - t_2 = \frac{I_o L}{V_{\text{in}}}(1 - \cos\alpha) \tag{6.69}$$

Therefore, there will be no power transfer and no charge or discharge interval when the switch turns on again in the next mode.

Mode IV **($t_3 \le t < t_4$)** In mode IV, as shown in Fig. 6.26(*e*), the inductor current is trapped and held constant at $i_L = I_o$, with $v_c = 0$:

$$i_L = I_o$$
$$v_c = 0$$

Mode IV will continue as long as the switch is on. By turning off the switch at $t = t_4 = T_s$, the switching cycle repeats. The dead time $t_4 - t_3$ is given by

$$t_4 - t_3 = T_s - t_1 - (t_2 - t_1) - (t_3 - t_2) \tag{6.70}$$

Voltage Gain We follow the same approach as in the ZCS case by using the energy balance concept. The input energy is given by

$$E_{\text{in}} = \int_0^{T_s} i_{\text{in}} V_{\text{in}}\, dt$$

i_{in} is the current, which is equal to $i_L(t)$. Hence, we have

$$E_{\text{in}} = \int_0^{t_1} i_L(t) V_{\text{in}}\, dt + \int_{t_1}^{t_2} i_L(t) V_{\text{in}}\, dt + \int_{t_2}^{t_3} i_L(t) V_{\text{in}}\, dt + \int_{t_3}^{T_s} i_L(t) V_{\text{in}}\, dt \tag{6.71}$$

The inductor current equals the output current in modes I and IV, and for $t_1 \le t < t_2$ and $t_2 \le t < t_3$, i_L is given in Eqs. (6.63) and (6.68), respectively. Substitute the inductor currents into Eq. (6.71) to yield Eq. (6.72):

$$\begin{aligned} E_{\text{in}} = V_{\text{in}}[I_o t_1 + I_o\sqrt{LC}\sin\omega_o(t_2 - t_1) + \frac{V_{\text{in}}}{2L}(t_3 - t_2)^2 \\ + I_o(t_3 - t_2)\cos\alpha + I_o(T_s - t_3)] \end{aligned} \tag{6.72}$$

Substituting for the time intervals t_1, $(t_2 - t_1)$, $(t_3 - t_2)$, and $(T_s - t_3)$ from Eqs. (6.62), (6.65), (6.69), and (6.70), respectively, and using the normalized parameters M, Q, and ω_o, Eq. (6.72) becomes

$$E_{\text{in}} = V_{\text{in}} I_o\left[\frac{-Q}{M\omega_o} - \frac{ML}{2R_o} + T_s - \frac{\alpha}{\omega_o} + \frac{ML}{R_o}\cos\alpha - \frac{ML}{2R_o}\cos^2\alpha\right] \tag{6.73}$$

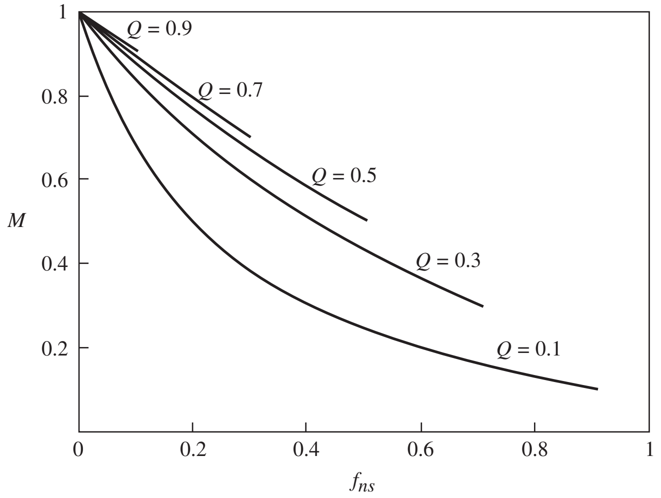


Figure 6.28 Control characteristic curve of M vs. f_{ns} for ZVS buck converter.

The output energy is expressed by

$$E_o = \int_0^{T_s} I_o V_o \, dt = I_o V_o T_s \tag{6.74}$$

Equating the input and output energy in Eqs. (6.73) and (6.74), the voltage gain expression becomes

$$M = 1 - \frac{f_{ns}}{2\pi}\left[\frac{M}{2Q} + \alpha + \frac{M}{Q}(1 - \cos\alpha)\right] \tag{6.75}$$

A plot of the control characteristic curve of M vs. f_{ns} is shown in Fig. 6.28.

The Boost Converter

In this section, we consider the quasi-resonant boost converter using the M-type switch as shown in Fig. 6.29(*a*), with its simplified circuit shown in Fig. 6.29(*b*). The four circuit modes of operation are shown in Fig. 6.30.

Mode I **($0 \le t < t_1$)** Assume for $t < 0$, the switch is closed while D is open. At $t = 0$, the switch is turned off, allowing the capacitor to charge by the constant current I_{in} as given by

$$I_{\text{in}} = i_L(t) = i_c(t) = C\frac{dv_c}{dt} \tag{6.76}$$

Since the initial capacitor voltage equals zero, Eq. (6.76) gives the following expression for $v_c(t)$:

$$v_c(t) = \frac{I_{\text{in}}}{C}t \tag{6.77}$$

The capacitor voltage reaches the output voltage at $t = t_1$, i.e., $v_c(t_1) = V_o$, resulting in

$$t_1 = \frac{CV_o}{I_{\text{in}}} \tag{6.78}$$

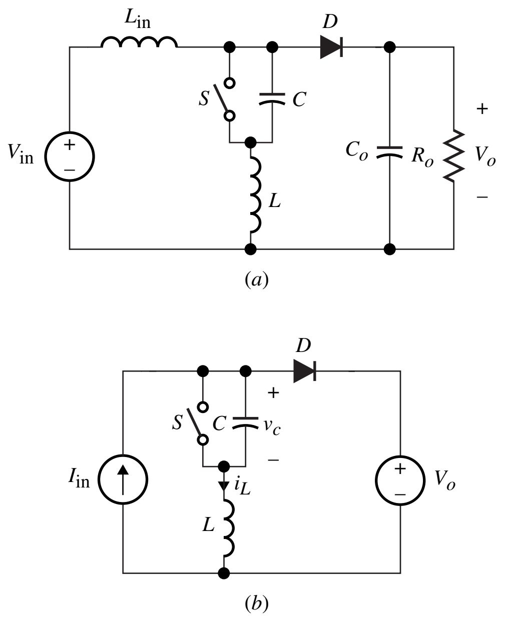


Figure 6.29 (*a*) Quasi-resonant boost converter with M-type switch. (*b*) Equivalent circuit.

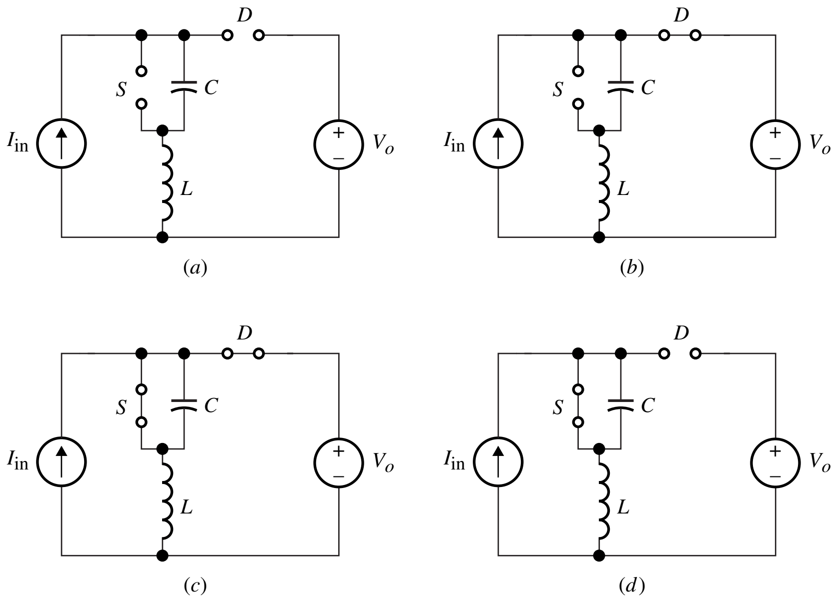


Figure 6.30 Equivalent circuit modes. (*a*) Mode I. (*b*) Mode II. (*c*) Mode III. (*d*) Mode IV.

At $t = t_1$, the diode starts conducting since $v_c = V_o$, and the converter enters mode II.

Mode II **($t_1 \le t < t_2$)** At $t = t_1$, the resonant stage begins since D is on and S is off, as shown in Fig. 6.30(*b*). When the capacitor voltage reaches the output voltage, i_L reaches the negative peak. The initial conditions are $v_c(t_1) = V_o$ and $i_L(t_1) = I_{in}$.

The expression for $v_c(t)$ is given by

$$v_c(t) = V_o + I_{in} Z_o \sin \omega_o (t - t_1) \tag{6.79}$$

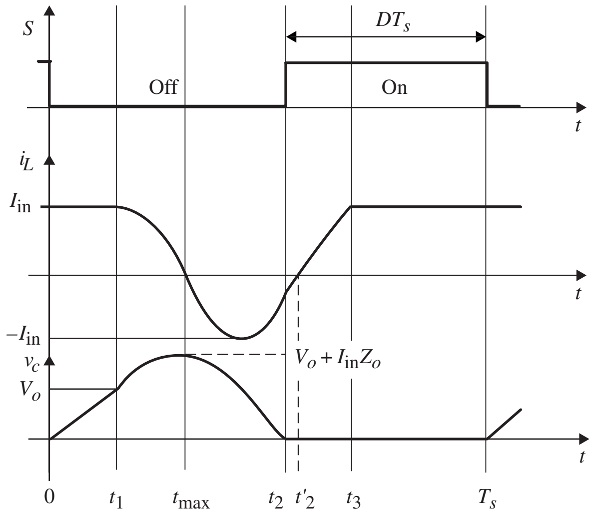


Figure 6.31 Steady-state waveforms for ZVS boost converter.

and the inductor current is

$$i_L(t) = I_{in}[1 + \cos\omega_o(t - t_1)] \tag{6.80}$$

Evaluating Eq. (6.79) at $t = t_2$ with $v_c(t_2) = 0$, the time interval between t_1 and t_2 can be found to be

$$(t_2 - t_1) = \frac{1}{\omega_o}\sin^{-1}\left(\frac{-V_o}{I_{in}Z_o}\right) \tag{6.81}$$

Mode III $(t_2 \le t < t_3)$ Mode III starts at t_2, when v_c reaches zero and S turns on at ZVS. The switch and the diode are both conducting, and the inductor current linearly increases to I_{in} as shown in Fig. 6.31. At $t = t_3$, the diode (anti-parallel diode) turns on, clamping the voltage across C to zero.

The initial conditions at $t = t_2$ are

$$V_c(t_2) = 0 \tag{6.82a}$$

$$I_L(t_2) = I_{in}(1 + \cos\omega_o(t_2 - t_1)) \tag{6.82b}$$

Because the capacitor voltage is zero, the inductor voltage is equal to the output voltage.

$$L\frac{di_L}{dt} = V_o \tag{6.83}$$

By integrating Eq. (6.83), the inductor current becomes

$$i_L(t) = \frac{V_o}{L}(t - t_2) + I_L(t_2) \tag{6.84}$$

To achieve ZVS, the switch can be turned on anytime after t_2 and before t_2'. At $t = t_2'$, the inductor current reverses polarity and the switch picks up the current.

At $t = t_3$, i_L reaches zero, resulting in the time interval given in Eq. (6.85):

$$(t_3 - t_2) = \frac{L}{V_o}i_L(t_2) \tag{6.85}$$

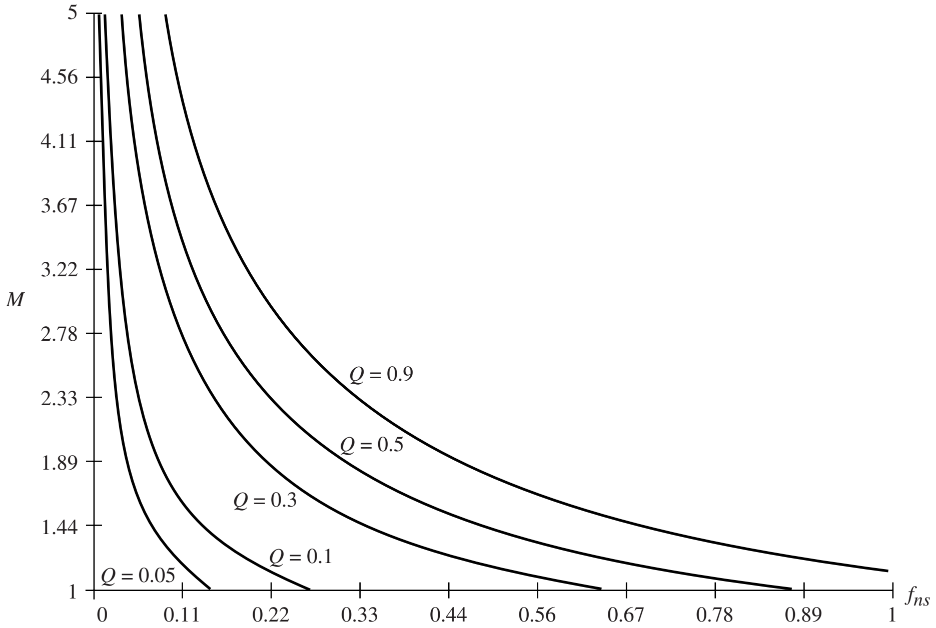


Figure 6.32 Control characteristic curve of M vs. f_{ns} for ZVT boost converter.

Substituting the initial condition into the equation, we get

$$(t_3 - t_2) = \frac{L}{V_o} I_{\text{in}}(1 + \cos\omega_o(t_2 - t_1)) \tag{6.86}$$

At $t = t_3$, the output diode turns off and the entire I_{in} current flows in the transistor and the inductor.

***Mode IV* ($t_3 \le t < t_4$)** At time t_3, the inductor current reaches zero and the output diode turns off, but the switch remains closed. The cycle repeats at $t = T_s$.

Voltage Gain As before, we use the conservation of energy per switching cycle to express the voltage gain. It can be shown that the voltage gain in terms of the normalized parameter is given in Eq. (6.87). A plot of the control characteristic curve of M vs. f_{ns} is shown in Fig. 6.32.

$$\frac{1-M}{M} = \frac{f_{ns}}{2\pi}\left[\frac{M}{2Q} + \alpha + \frac{M}{Q}(1 - \cos\alpha)\right] \tag{6.87}$$

EXAMPLE 6.5

Design a ZVS quasi-resonant boost converter for the following design parameters: $V_{\text{in}} = 30$ V, $P_o = 30$ W at $V_o = 38$ V, $f_{ns} = 0.4$, and $T_s = 4\ \mu\text{s}$. Assume the output voltage ripple is limited to 2% at $D = 0.4$.

SOLUTION The voltage gain is $M = V_o/V_{\text{in}} = 1.3$, and with $f_{ns} = 0.4$, we obtain $Q = 0.2$.

Using the switching frequency $f_s = 1/T_s = 1/4\ \mu\text{s} = 250$ kHz, and $f_o = f_s/0.4 = 250/0.4 = 625$ kHz, the resonant frequency is obtained from

$$\omega_o = \frac{1}{\sqrt{LC}} = (2\pi)(625) \times 10^3 \text{ rad/s}$$

The second equation in terms of L and C is obtained from

$$Q = \frac{R_o}{Z_o} = \frac{R_o}{\sqrt{L/C}} = 0.2$$

The load resistance is

$$R_o = \frac{38^2}{30} = 48.13\ \Omega$$

Substituting in the above relation for Q, we obtain

$$\sqrt{\frac{L}{C}} = \frac{48.13}{0.2} = 240.65\ \Omega$$

Solving the above two equations for C and L, we obtain

$$C = \frac{1}{(2\pi)(625)(10^3)(240.65)} = 1.06\ \text{nF}$$

$$L = 1.06 \times 10^{-9} \times (240.65)^2 = 61.3\ \mu\text{H}$$

To calculate L_o and C_o, using a voltage ripple of 2%, we use the following relation:

$$\frac{D}{f_s R_o C_o} = 0.02$$

where

$$D = 0.4$$

$$C_o = \frac{0.4}{f_s R_o}\frac{100}{0.2}$$

$$= \frac{40}{0.2 \times 250 \times 10^3 \times 48.13} = 16.6\ \mu\text{F}$$

The critical inductor value is given by

$$L_{\text{crit}} = \frac{R_o}{2f_s}(1-D)^2 D$$

$$= \frac{48.13}{2 \times 250 \times 10^3}(1-0.4)^2(0.4)$$

$$= 13.9\ \mu\text{H}$$

To achieve a limited ripple current, it is recommended that L_o be set at about 100 times the critical inductor value. So we select $L_o = 1.4$ mH.

EXERCISE 6.5

The quasi-resonant boost converter using the L-type switch is shown in Fig. 6.33(*a*), and its simplified circuit is shown in Fig. 6.33(*b*). Derive the expression for the voltage gain.

The Buck-Boost Converter

The ZVS buck-boost converter with an M-type switch is shown in Fig. 6.34(*a*), with its equivalent circuit shown in Fig. 6.34(*b*).

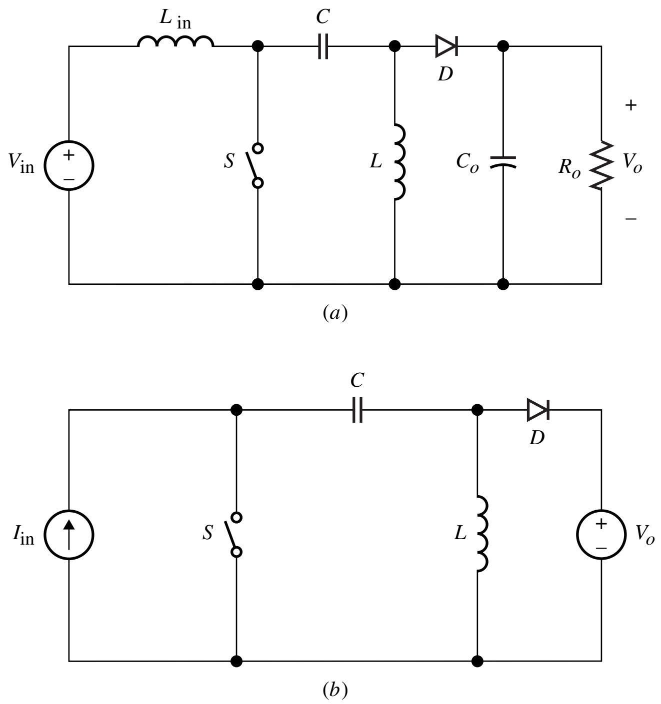


Figure 6.33 (*a*) Quasi-resonant boost converter with L-type switch. (*b*) Simplified equivalent circuit.

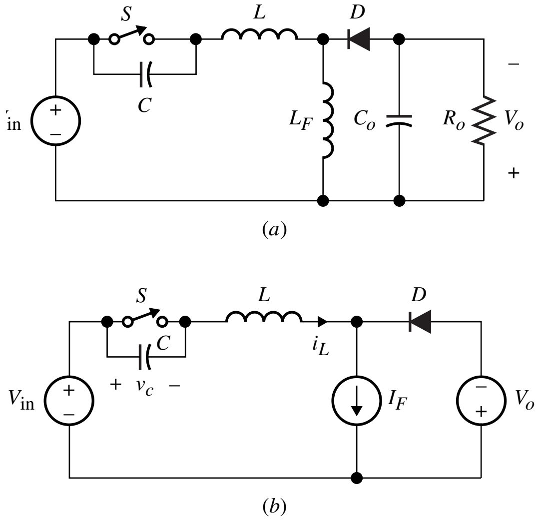


Figure 6.34 (*a*) ZVS buck-boost converter with M-type switch. (*b*) Simplified equivalent circuit.

Like the buck and the boost converters, the buck-boost converter steady-state operation leads to four modes of operation as shown in Fig. 6.35. Figure 6.35(*e*) shows the typical steady-state waveforms for v_c and i_L.

Following a similar analysis as before, it can be shown that the voltage gain in terms of M, Q, and f_{ns} is given in Eq. (6.88).

$$M = \frac{1}{\dfrac{f_{ns}}{2\pi}\left[\alpha + \dfrac{M}{2Q} + \dfrac{M}{Q}(1 - \cos\alpha)\right]} - 1 \qquad (6.88)$$

Figure 6.36 shows the control characteristic curve for M vs. f_{ns}.

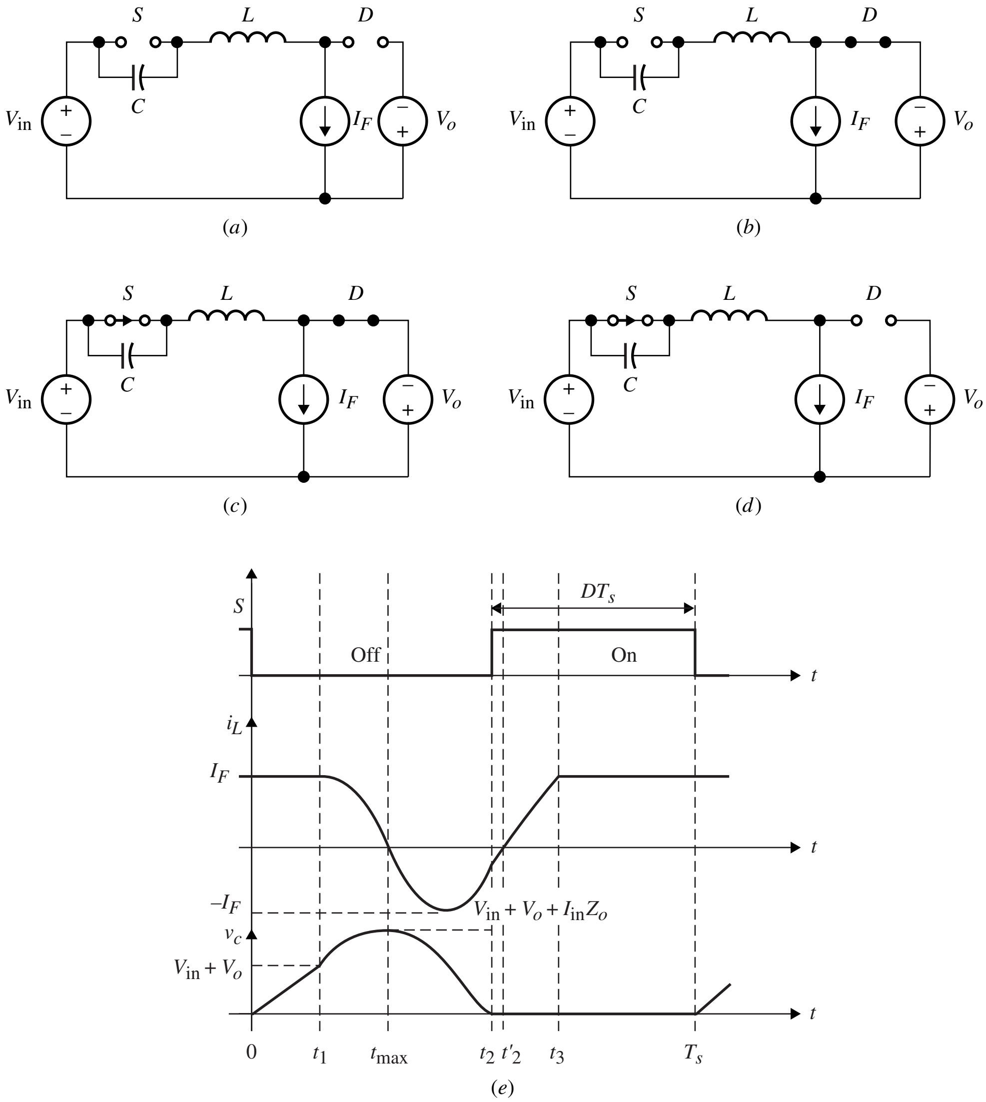


Figure 6.35 Equivalent circuits for (*a*) mode I, (*b*) mode II, (*c*) mode III, and (*d*) mode IV. (*e*) Steady-state waveforms for v_c and i_L.

6.6 GENERALIZED ANALYSIS FOR ZCS[2]

It can be shown from the preceding analysis of the quasi-resonant ZCS and ZVS PWM converters that all dc-dc converter families share the same switching network, with the orientation depending on the topology type (buck, boost, buck-boost, Cuk, Zeta, SEPIC, etc.). As a result, the switching network representation and analysis, including the switching waveforms, can be generalized for each converter family.

The generalized analysis means that in order to derive the analysis and the design curves for a family of converters, only the generalized switching cell with the generalized and normalized parameters for that family need to be analyzed. The equations for the

[2]This section can be skipped without loss of continuity.

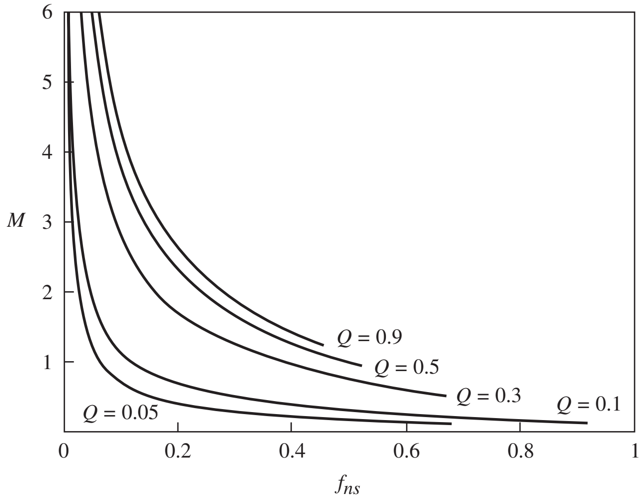


Figure 6.36 Control characteristic curve of M vs. f_{ns} for ZVS buck-boost converter.

generalized cell will be the generalized equations that describe any converter that uses this specific cell. By using generalized parameters, it is possible to generate a single *transformation table* from which the voltage ratios and other important design parameters for each converter can be obtained directly.

6.6.1 The Generalized Switching Cell

Figure 6.37(*a*) and (*b*) shows the generalized switching cells of the quasi-resonant PWM ZCS and ZVS converters, respectively. Note that these cells are all of the common-

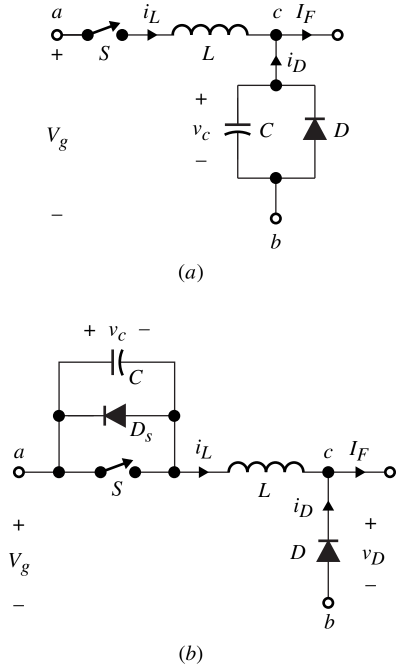


Figure 6.37 Switching cells. (*a*) ZCS quasi-resonant converter cell. (*b*) ZVS quasi-resonant converter cell.

ground, three-terminal, two-port network type. The following parameters, will be used throughout this discussion:

- The normalized cell input voltage, V_{ng}:

$$V_{ng} = \frac{V_g}{V_{\text{in}}}$$

  where V_g is the switching-cell average input voltage as shown in Fig. 6.37.
- The normalized cell output current, I_{nF}:

$$I_{nF} = \frac{I_F}{I_o}$$

  where I_F is the switching-cell average output current.
- The normalized filter capacitor voltage, V_{nF}:

$$V_{nF} = \frac{V_F}{V_{\text{in}}}$$

  where V_F is the filter capacitor average voltage.
- The normalized filter inductor current, I_{nT}:

$$I_{nT} = \frac{I_T}{I_o}$$

  where I_T is the filter inductor average current (in the ZCS QSW CC family).
- The normalized cell output average voltage, V_{nbc}:

$$V_{nbc} = \frac{V_{bc}}{V_{\text{in}}}$$

  where V_{bc} is the switching-cell average output voltage.
- The normalized current entering node *b* in the switching cell, I_{nb}:

$$I_{nb} = \frac{I_b}{I_o}$$

  where I_b is the average current entering node *b*.

Note that the generalized transformation table that will be presented next includes the generalized parameters V_{ng}, I_{nF}, V_{nbc}, I_{nb}, V_{nF}, and I_{nT}, which are the normalized versions of the parameters V_g, I_F, V_{bc}, I_b, V_F, and I_T. It will be noted that I_{nb}, V_{nF}, and I_{nT} have the same normalized quantity, as do V_{ng} and I_{nF}.

6.6.2 The Generalized Transformation Table

The derivation of the generalized transformation table for the converter families is beyond the scope of this book. The generalized transformation table is shown in Table 6.1.

By applying the appropriate cell to the conventional dc-dc converters, the ZCS quasi-resonant converter (QRC) family can be formed as shown in Fig. 6.38.

In the next section, the modes of operation for the ZVS QRC switching network will be discussed briefly, and the main switching waveforms will be drawn in terms of the generalized parameters.

Table 6.1 Generalized Transformation Table

	V_{ng}, I_{nF}	V_{nF}, I_{nT}, I_{nb}	V_{nbc}
Buck	1	$1 - M$	$-M$
Boost	M	1	$1 - M$
Buck-boost, Cuk, Zeta, and SEPIC	$1 + M$	1	$-M$

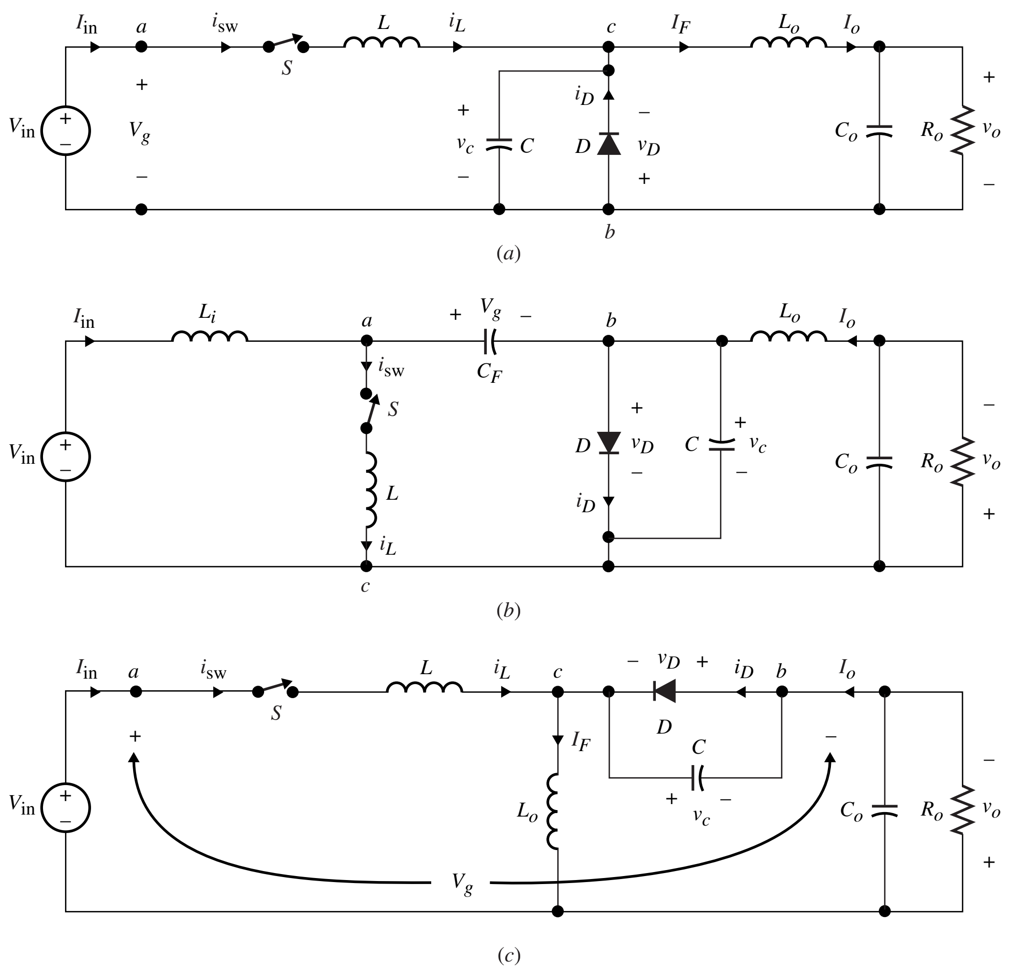


Figure 6.38 The dc-dc ZCS QRC family. (*a*) Buck. (*b*) Boost. (*c*) Buck-boost.

6.6.3 Basic Operation of the ZCS QRC Cell

Typical switching waveforms for the cell in Fig. 6.37(*a*) are shown in Fig. 6.39. Table 6.2 shows the conditions of the switches and diodes in each mode. It can be shown that there are four modes of operation.

Mode 1 **($t_0 \leq t < t_1$)** It is assumed that before $t = t_0$, S was off and D was on in order to carry I_F. The resonant inductor L was carrying no current, and the resonant capacitor C voltage was zero. Mode 1 starts when S is turned on while D is on, which

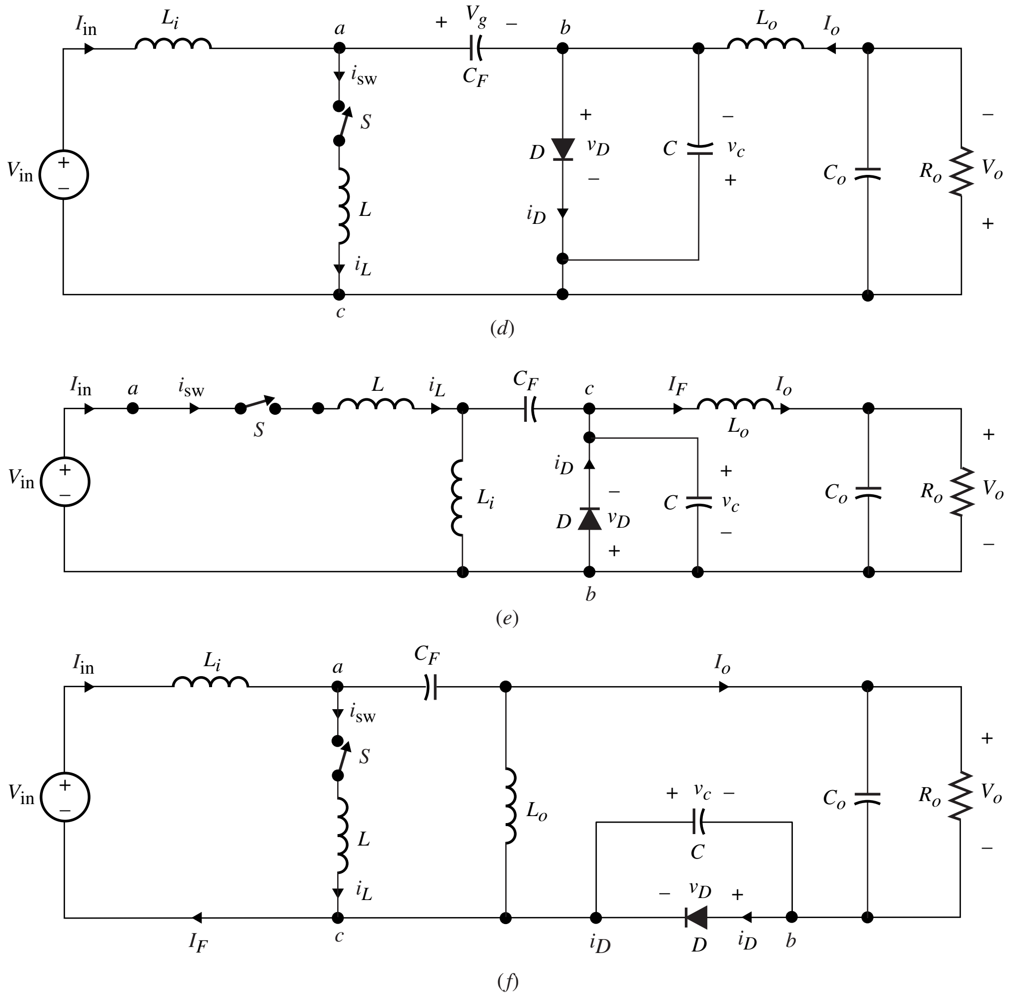


Figure 6.38 The dc-dc ZCS QRC family. (*d*) Cuk. (*e*) Zeta. (*f*) SEPIC (*continued*).

causes L to charge up linearly until the current through it becomes equal to I_F at $t = t_1$, causing D to turn off.

Mode 2 $(t_1 \leq t < t_2)$ Mode 2 starts when D turns off while S is on, initiating a resonant stage between C and L until the current through L drops to zero at $t = t_2$, causing S to turn off at zero current (soft switching).

Mode 3 $(t_2 \leq t < t_3)$ Mode 3 starts when S turns off at zero current. The resonant capacitor starts discharging linearly, causing the voltage across it to drop to zero again, in turn causing D to turn on at zero voltage at $t = t_3$.

Mode 4 $(t_3 \leq t < t_0 + T_s)$ Mode 4 is a steady-state mode, and nothing happens in it until S is turned on again to start the next switching cycle.

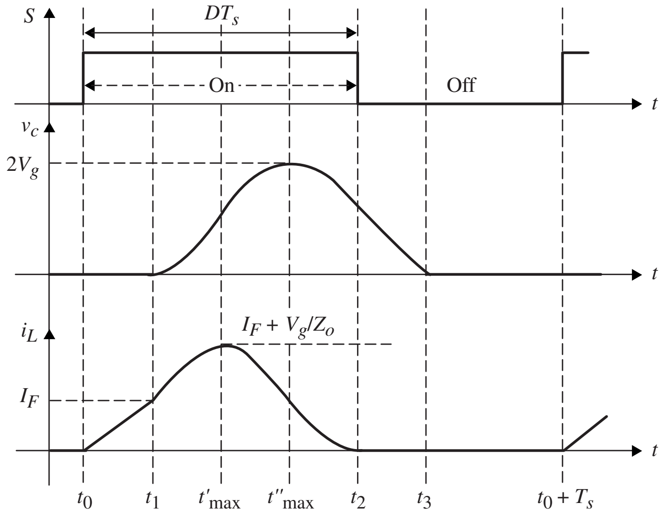


Figure 6.39 Main ZCS QRC switching-cell waveforms.

Table 6.2 Switch and Diode Conditions

	S	D
Mode 1 $(t_0 \le t < t_1)$	On	On
Mode 2 $(t_1 \le t < t_2)$	On	Off
Mode 3 $(t_2 \le t < t_3)$	Off	Off
Mode 4 $(t_3 \le t < t_0 + T_s)$	Off	On

Generalized Steady-State Analysis

From the description of the modes of operation, the equivalent circuit for each mode can be drawn as in Fig. 6.40. These equivalent circuits can be used along with the description of the modes to write the mathematical equations for each mode as follows (knowing that $v_c(t_0) = 0$ and $i_L(t_0) = 0$):

Mode 1 $(t_0 \le t < t_1)$

$$v_c(t) = 0$$
$$i_L(t) = \frac{V_g}{L}(t - t_0) \tag{6.89}$$

with the initial conditions at $t = t_1$ given by

$$v_c(t_1) = 0$$
$$i_L(t_1) = I_F$$

Mode 2 $(t_1 \le t < t_2)$

$$v_c(t) = V_g[1 - \cos \omega_o(t - t_1)] \tag{6.90}$$

$$i_L(t) = I_F + \frac{V_g}{Z_o}\sin \omega_o(t - t_1) \tag{6.91}$$

At $t = t_2$, we have $i_L(t_2) = 0$.

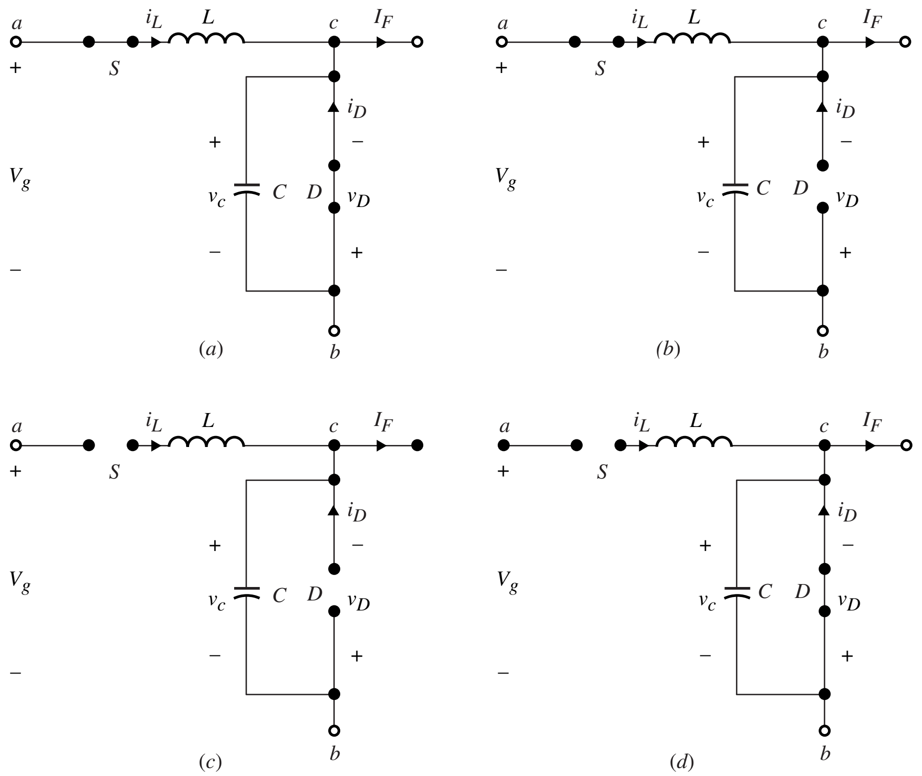


Figure 6.40 Equivalent circuits for (*a*) mode 1, (*b*) mode 2, (*c*) mode 3, and (*d*) mode 4.

Mode 3 $(t_2 \le t < t_3)$

$$v_c(t) = -\frac{I_F}{C}(t - t_2) + V_g(1 - \cos\beta) \tag{6.92}$$

$$i_L(t) = 0 \tag{6.93}$$

At $t = t_3$, we have $v_c(t_3) = 0$.

Mode 4 $(t_3 \le t < t_0 + T_s)$

$$v_c(t) = 0 \tag{6.94}$$

$$i_L(t) = 0 \tag{6.95}$$

Generalized Interval Equations

To simplify the analysis, the following time intervals are defined:

$$\alpha = \omega_o(t_1 - t_0)$$
$$\beta = \omega_o(t_2 - t_1)$$
$$\gamma = \omega_o(t_3 - t_2)$$
$$\delta = \omega_o((t_0 + T_s) - t_3)$$

These intervals can be derived as follows:

- From Eq. (6.89), α is given by

$$\alpha = \omega_o(t_1 - t_0) = \frac{Z_o I_F}{V_g}$$

  Using the normalized parameters, we have

$$\alpha = \omega_o(t_1 - t_0) = \frac{M I_{nF}}{Q V_{ng}} \tag{6.96}$$

- From Eq. (6.91), β is given by

$$\beta = \omega_o(t_2 - t_1) = \sin^{-1}\left(-\frac{M I_{nF}}{Q V_{ng}}\right) \tag{6.97}$$

- From Eq. (6.92), γ is given by

$$\gamma = \omega_o(t_3 - t_2) = \frac{Q V_{ng}}{M I_{nF}}(1 - \cos\beta)$$

- From Fig. 6.39 and the intervals α, β, and γ, we have δ given by

$$\delta = \omega_o((t_0 + T_s) - t_3) = \frac{2\pi}{f_{ns}} - \alpha - \beta - \gamma \tag{6.98}$$

Generalized Gain Equation

The cell output-to-input generalized gain can be found using the average output diode D voltage as follows:

$$\begin{aligned} V_{D,\text{ave}} &= -V_{c,\text{ave}} \\ &= -\frac{1}{T_s}\int_{t_0}^{t_0+T_s} v_c(t)\,dt \\ &= -\frac{1}{T_s}\left[V_g\left((t_2 - t_1) - \frac{\sin\beta}{\omega_o}\right) - \frac{I_F}{2C}(t_3 - t_2)^2 + V_g(1 - \cos\beta)(t_3 - t_2)\right] \end{aligned}$$

By using the normalized parameters defined previously, we have

$$V_{nD} = \frac{f_{ns}}{2\pi}\left[\frac{M I_{nF}}{2Q}\gamma^2 - V_{ng}(\beta + \gamma - \sin\beta - \gamma\cos\beta)\right] \tag{6.99}$$

By substituting for the generalized parameters (V_{nD}, V_{ng}, and I_{nF}) from Table 6.1, we arrive at the gain equation for each converter in the family.

Generalized ZCS Condition

It can be noted from Fig. 6.39 that S can be turned off at any time after $t = t_2$. The generalized condition to achieve zero-current switching can be expressed as follows:

$$\frac{2\pi}{f_{ns}} D \geq \alpha + \beta \tag{6.100}$$

Generalized Peak Resonant Inductor Current (Peak Switch Current)

The peak resonant inductor current or peak switch current occurs at $t = t'_{\max}$, when $\omega_o(t'_{\max} - t_1) = \pi/2$. Using Eq. (6.92) at $t = t'_{\max}$,

$$I_{nL\text{-peak}} = I_{nF} + \frac{QV_{ng}}{M} \tag{6.101}$$

The peak resonant capacitor voltage or peak diode voltage occurs at $t = t''_{\max}$, when $\omega_o(t''_{\max} - t_1) = \pi$. Using Eq. (6.91) at $t = t'_{\max}$,

$$V_{nL\text{-peak}} = 2V_{ng}$$

Design Control Curves

By substituting the generalized parameters from Table 6.1 in the generalized equations, design equations for each topology in the family can be found and their design curves can be plotted.

Figure 6.41 shows the control characteristic curves of M vs. f_{ns} for the ZCS QRC family as an example. Figure 6.42 shows the average and the rms switch currents as a function of the voltage gain.

6.6.4 Basic Operation of the ZVS QRC Cell

Figure 6.37(*b*) shows the generalized ZVS QRC switching cell, which is formed by adding the resonant capacitor C (which can be considered the switch's internal capacitor) and the resonant inductor L to the conventional switching cell. As discussed before, this is also known as an M-type resonant switch.

By applying this cell to the conventional dc-dc converters in Fig. 6.7, the ZVS QRC family can be formed as shown in Fig. 6.43. In this section, the modes of operation for the ZVS QRC switching network will be discussed briefly and the main switching waveforms will be drawn.

The typical switching waveforms for the ZVS QRC cell are shown in Fig. 6.44. Table 6.3 shows the conduction status of the switches and diodes in each mode. As before, there are four modes of operation. It is assumed that before $t = t_0$, S was on and D was off. The resonant inductor L was carrying a current equal to I_F, and the resonant capacitor C voltage was zero.

Mode 1 **($t_0 \le t < t_1$)** Mode 1 starts when S is turned on while D is off, which causes C to charge up linearly until its voltage reaches a value equal to V_g at $t = t_1$, causing D to start conducting. The resonant inductor current during this mode doesn't change and is equal to I_F.

Mode 2 **($t_1 \le t < t_2$)** Mode 2 starts when D turns on while S is off, causing a resonant stage between C and L until the voltage across C tends to go negative, forcing the switch diode D_s to turn on at $t = t_2$. After this time, S can be turned on at zero voltage.

Mode 3 **($t_2 \le t < t_3$)** Mode 3 starts when S is turned on at zero voltage (i.e., zero-voltage switching). The resonant inductor current starts charging up linearly until it reaches I_F, causing D to turn off at zero current at $t = t_3$.

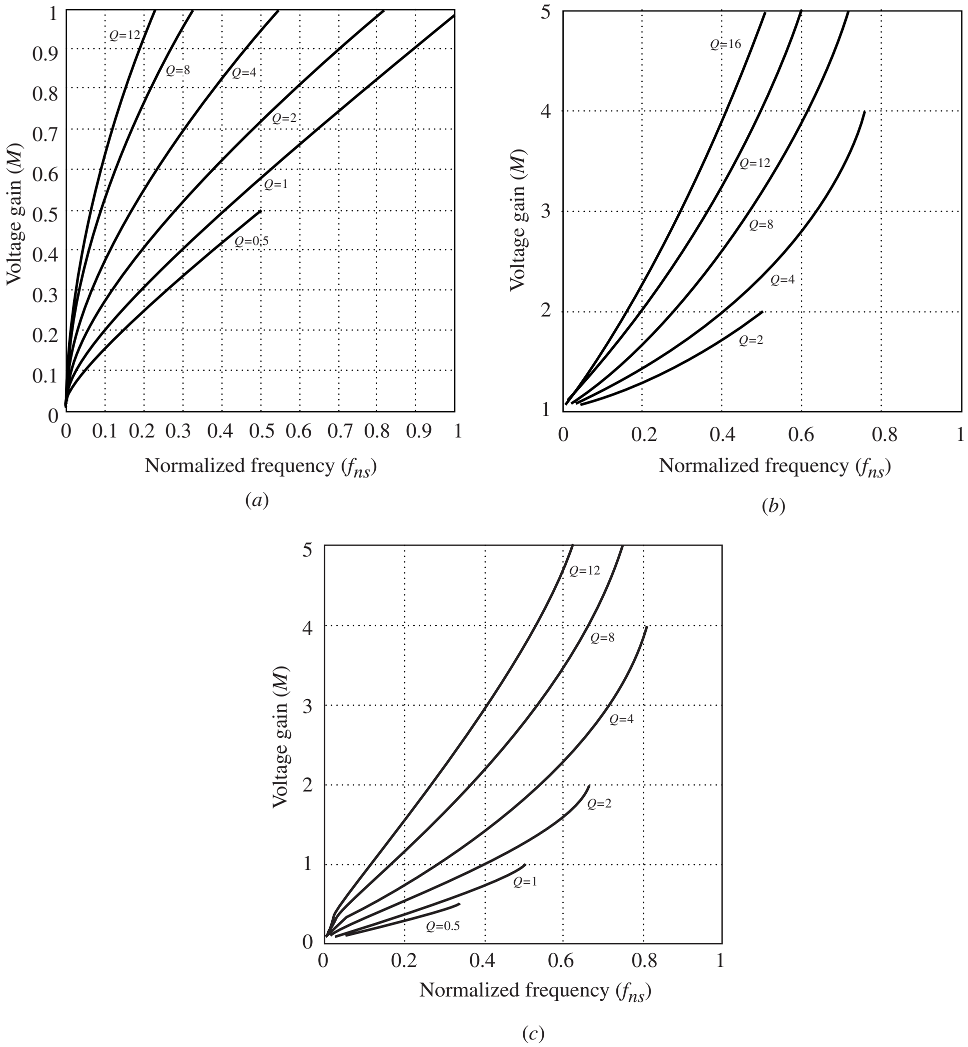


Figure 6.41 Control characteristic curves of M vs. f_{ns} for (*a*) ZCS QRC buck, (*b*) ZCS QRC boost, (*c*) ZCS QRC buck-boost, Cuk, Zeta, and SEPIC.

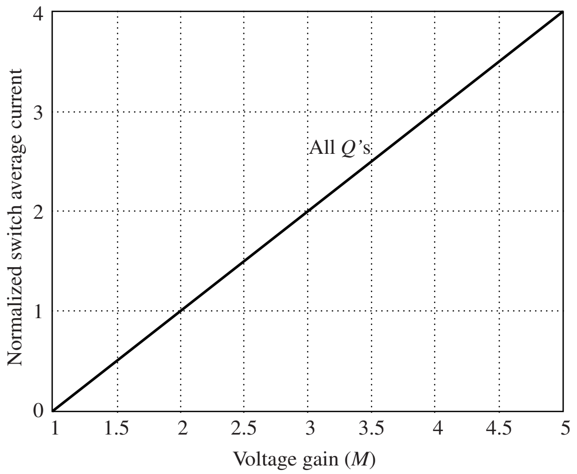
Mode 4 **($t_3 \leq t < t_0 + T_s$)** Mode 4 is a steady-state mode, and nothing happens in it until S is turned off again to start the next switching cycle.

Generalized Steady-State Analysis

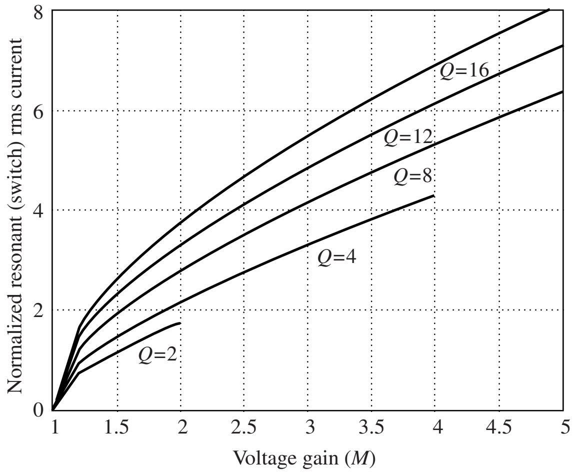
From the description of the modes of operation, the equivalent circuit for each mode can be drawn as shown in Figure 6.45. These equivalent circuits can be used along with the description of the modes to write the mathematical equations for each mode as follows (knowing that $v_c(t_0) = 0$ and $i_L(t_0) = I_F$):

Mode 1 **($t_0 \leq t < t_1$)**

$$\begin{aligned} v_c(t) &= \frac{I_F}{C}(t - t_0) \\ i_L(t) &= I_F \end{aligned} \tag{6.102}$$



(a)



(b)

Figure 6.42 Some of the ZCS QRC boost main switch (S) normalized stresses.
(a) Normalized average current.
(b) Normalized rms current.

$$\begin{aligned} v_c(t_1) &= V_g \\ i_L(t_1) &= I_F \end{aligned} \tag{6.103}$$

Mode 2 $(t_1 \le t < t_2)$

$$\begin{aligned} v_c(t) &= V_g + Z_o I_F \sin \omega_o (t - t_1) \\ i_L(t) &= I_F \cos \omega_o (t - t_1) \end{aligned} \tag{6.104}$$

$$\begin{aligned} v_c(t_2) &= 0 \\ i_L(t_2) &= I_F \cos \omega_o (t_2 - t_1) = 0 \end{aligned} \tag{6.105}$$

Mode 3 $(t_2 \le t < t_3)$

$$\begin{aligned} v_c(t) &= 0 \\ i_L(t) &= I_F \cos \omega_o (t_2 - t_1) + \frac{V_g}{L}(t - t_2) \end{aligned} \tag{6.106}$$

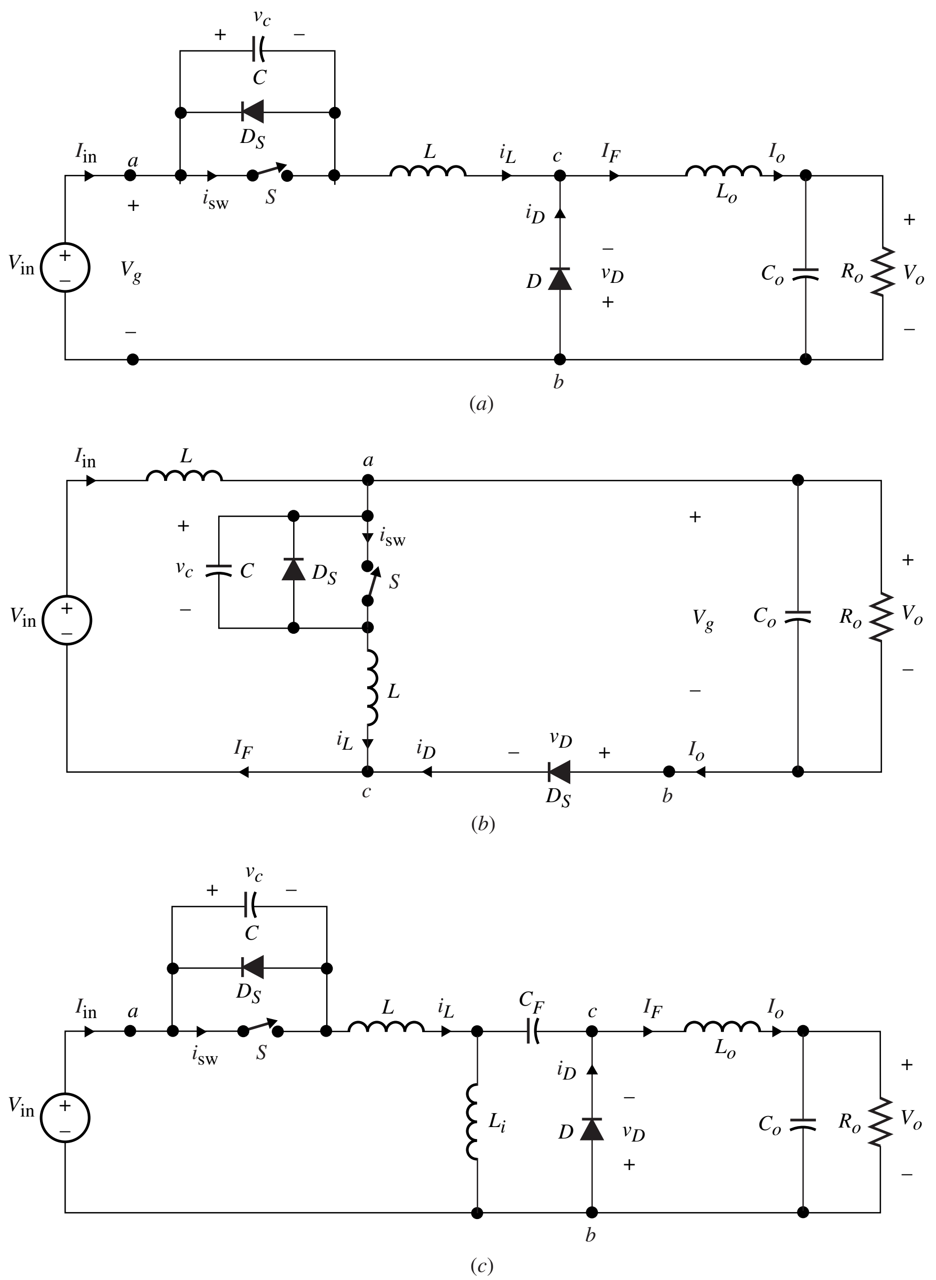


Figure 6.43 The dc-dc ZVS QRC family. (*a*) Buck. (*b*) Boost. (*c*) Buck-boost.

$$
\begin{aligned}
v_c(t_3) &= 0 \\
i_L(t_3) &= I_F
\end{aligned}
\tag{6.107}
$$

Mode 4 **($t_3 \leq t < t_0 + T_s$)**

$$
\begin{aligned}
v_c(t) &= 0 \\
i_L(t) &= I_F
\end{aligned}
$$

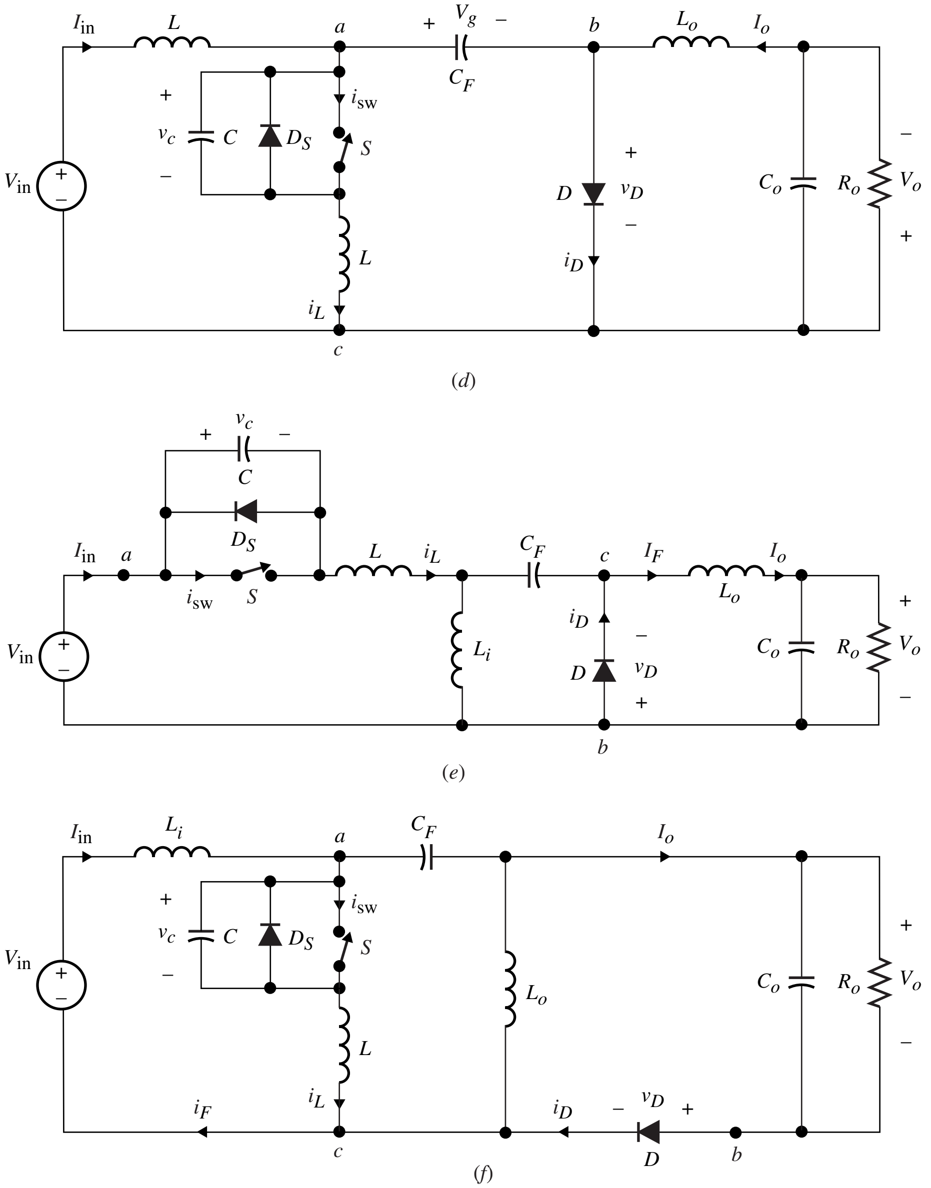


Figure 6.43 The dc-dc ZVS QRC family. (*d*) Cuk. (*e*) Zeta. (*f*) SEPIC (*continued*).

Generalized Interval Equations

To simplify the analysis, the following time intervals are defined:

$$\alpha = \omega_o(t_1 - t_0)$$
$$\beta = \omega_o(t_2 - t_1)$$
$$\gamma = \omega_o(t_3 - t_2)$$
$$\delta = \omega_o((t_0 + T_s) - t_3)$$

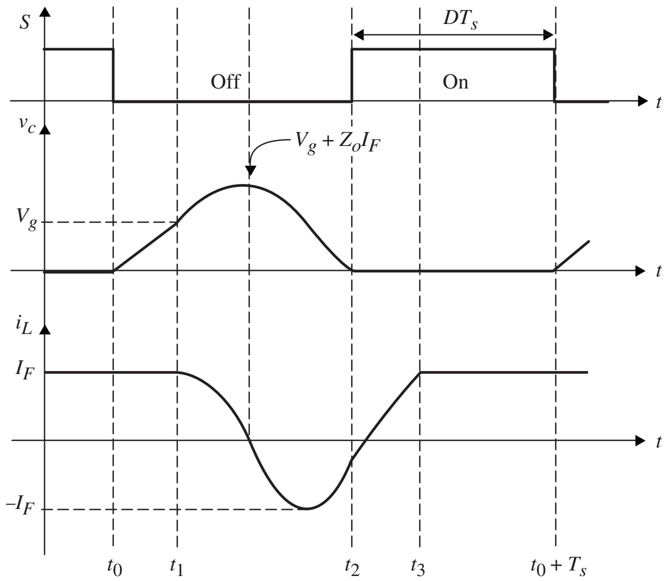


Figure 6.44 Main ZVS QRC switching-cell waveforms.

Table 6.3 Switches and Diode Conditions

	S	D	D_s
Mode 1 $(t_0 \le t < t_1)$	Off	Off	Off
Mode 2 $(t_1 \le t < t_2)$	Off	On	Off
Mode 3 $(t_2 \le t < t_3)$	On	On	On
Mode 4 $(t_3 \le t < t_0 + T_s)$	On	Off	Don't care

These intervals can be derived as follows:

- From Eqs. (6.102) and (6.103),

$$\alpha = \omega_o(t_1 - t_0) = \frac{V_g}{Z_o I_F}$$

  Using the normalized parameters defined earlier,

$$\alpha = \omega_o(t_1 - t_0) = \frac{QV_{ng}}{MI_{nF}}$$

- From Eqs. (6.104) and (6.105),

$$\beta = \omega_o(t_2 - t_1) = \sin^{-1}\left(-\frac{QV_{ng}}{MI_{nF}}\right)$$

- From Eqs. (6.106) and (6.107),

$$\gamma = \omega_o(t_3 - t_2) = \frac{MI_{nF}}{QV_{ng}}(1 - \cos\beta)$$

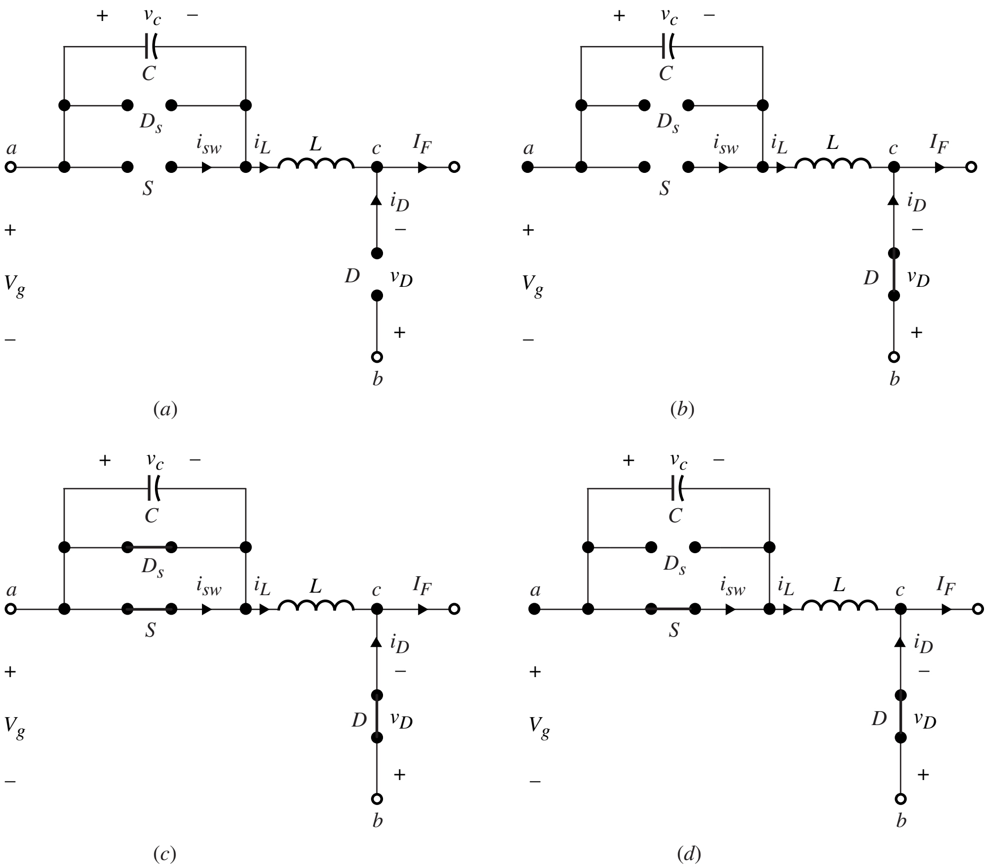


Figure 6.45 Equivalent circuits for (*a*) mode 1, (*b*) mode 2, (*c*) mode 3, and (*d*) mode 4.

- From Fig. 6.44 and the preceding intervals,

$$\delta = \omega_o((t_0 + T_s) - t_3) = \frac{2\pi}{f_{ns}} - \alpha - \beta - \gamma$$

Generalized Gain Equation

The cell output-to-input generalized gain can be found using the average output diode *D* voltage as follows:

$$\begin{aligned} V_{D,\,\text{ave}} &= V_{s,\,\text{ave}} - V_g \\ &= \left[\frac{1}{T_s}\int_{t_0}^{t_0+T_s} v_s(t)\,dt\right] - V_g \\ &= \frac{1}{T_s}\left[\frac{I_F}{2C}(t_1 - t_0)^2 + V_g(t_2 - t_1) - \frac{Z_o I_F}{\omega_o}(\cos\omega_o(t_2 - t_1) - 1)\right] - V_g \end{aligned}$$

By using the normalized parameters, we have

$$V_{nD} = \frac{f_{ns}}{2\pi}\left[\frac{MI_{nF}}{2Q}\alpha^2 + V_{ng}\beta + \frac{MI_{nF}}{Q}(1 - \cos\beta)\right] - V_{ng} \tag{6.108}$$

By substituting for the generalized parameters (V_{nD}, V_{ng}, and I_{nF}) from Table 6.1 in Eq. (6.108), we arrive at the gain equation for each converter in the family.

Generalized ZVS Condition

It can be noted from Fig. 6.43 that S must be turned on after $t = t_2$ and before $t = t_3$—in other words, before $i_L(t) = I_F$, which causes D to turn off ($i_D(t) = I_F - i_L(t)$) at $t = t_3$ causing $v_c(t)$ to start charging up linearly again. The generalized condition to achieve zero-voltage switching can be expressed as follows:

$$\alpha + \beta \le \frac{2\pi}{f_{ns}}(1 - D) \le \alpha + \beta + \gamma$$

Design Control Curves

By substituting for the generalized parameters from Table 6.1 in the generalized equations, design equations for each topology in the family can be found. Figure 6.46 shows

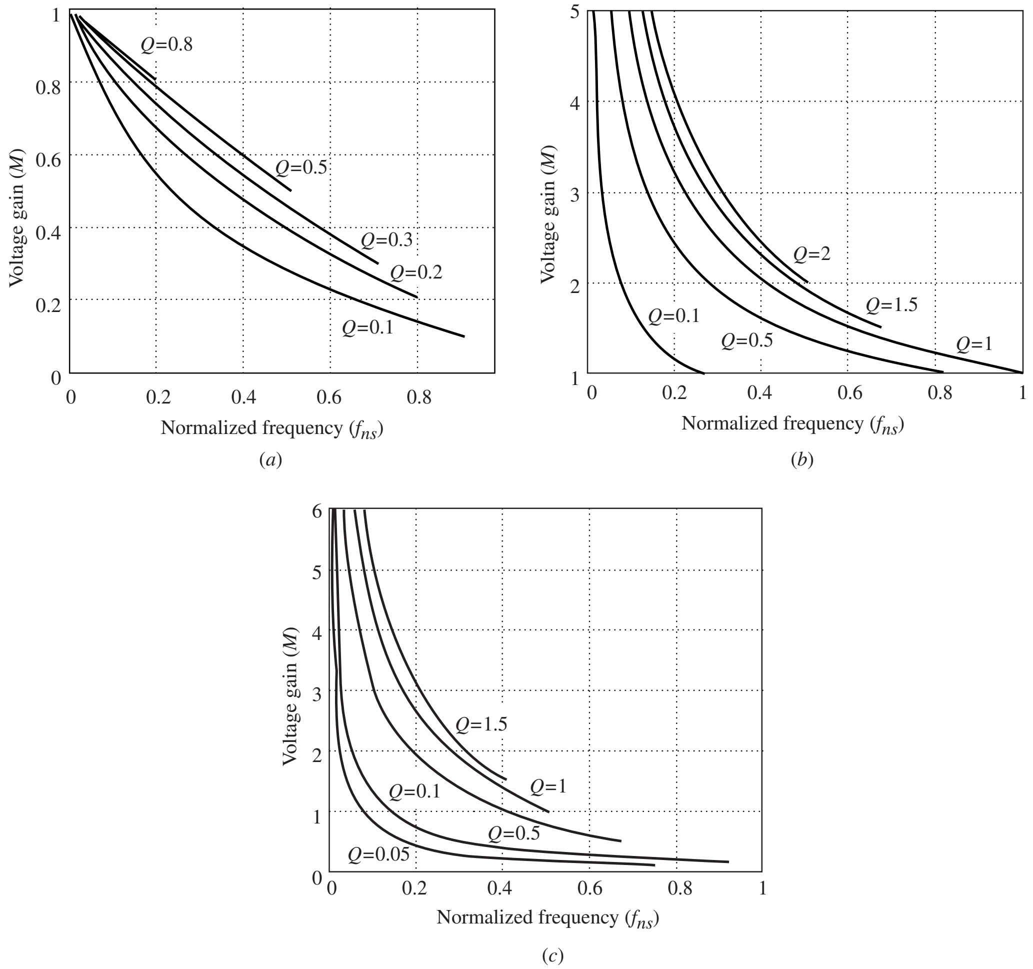


Figure 6.46 Control characteristic curves of M vs. f_{ns} for (*a*) ZVS QRC buck, (*b*) ZVS QRC boost, (*c*) ZVS QRC buck-boost, Cuk, Zeta, and SEPIC.

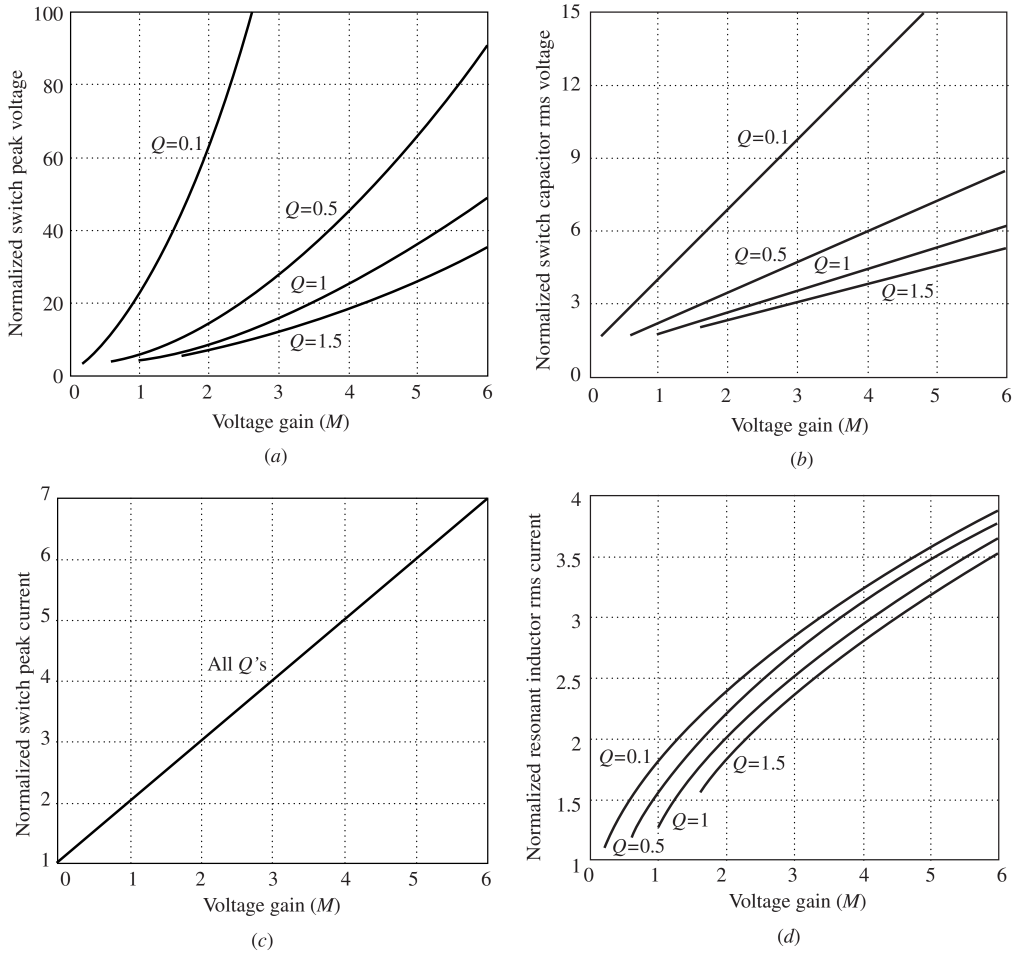


Figure 6.47 Some of the ZVS QRC buck-boost stresses. (*a*) Normalized switch peak voltage. (*b*) Normalized switch rms voltage. (*c*) Normalized switch peak current. (*d*) Normalized switch rms current.

the control characteristic curves for the ZVS QRC family. Many other curves can also be plotted; as an example, Fig. 6.47 shows the peak and the rms switch currents as a function of the voltage gain, *M*.

6.7 ZERO-VOLTAGE AND ZERO-CURRENT TRANSITION CONVERTERS

Traditional converters operate with a sinusoidal current through the power switches, which results in high peak and rms currents for the power transistors and high voltage stresses on the rectifier diodes. Furthermore, when the line voltage or load current varies over a wide range, quasi-resonant converters are modulated with a wide switching frequency range, making the circuit design difficult to optimize. As a compromise between the PWM and resonant techniques, various soft-switching PWM converter techniques have been proposed lately to combine the desirable features of the conventional PWM and quasi-resonant techniques without a significant increase in the circulating energy.

6.7.1 Switching Transition

To overcome the limitations of the quasi-resonant converters, zero-voltage transition (ZVT) or zero-current transition (ZCT) is the solution. Instead of using a series resonant network across the power switch, a shunt resonant network is used across the power switch. A partial resonance is created by the shunt resonant network to achieve ZCS or ZVS during the switching transition. This retains the advantages of a PWM converter because after the switching transition is over, the circuit reverts to the PWM operation mode.

The features of the ZCT PWM and ZVT PWM soft-switching converters are summarized as follows:

- Zero-current/voltage turn-off/on for the power switch
- Low voltage/current stresses of the power switch and rectifier diode
- Minimal circulating energy
- Constant-frequency operation
- Soft switching for a wide line and load range

One disadvantage is that the auxiliary switch does not operate with soft switching; it is hard-switching, but the switching loss is much lower than that of a PWM converter. Another disadvantage is that the transformer leakage inductance is not utilized, which is similar to the quasi-resonant converters. Therefore, the transformer should be designed with minimum leakage inductance.

The ZVT and ZCT converters differ from the conventional PWM converters by the introduction of a resonant branch, shown in Fig. 6.48. Figure 6.48(*a*) shows the

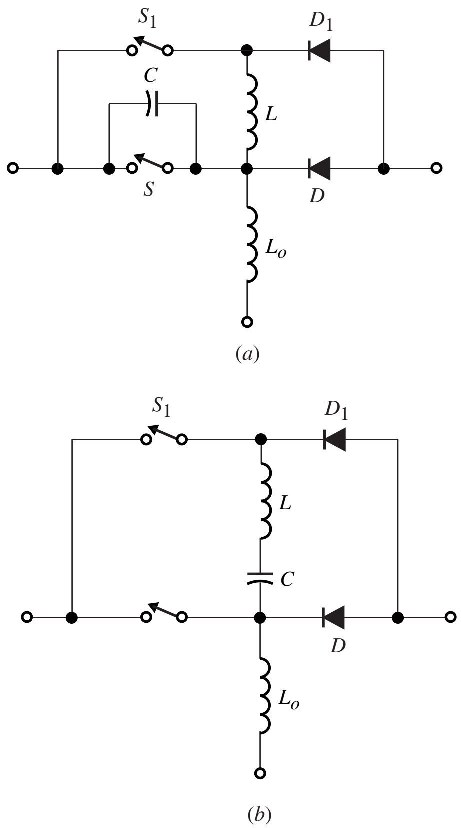


Figure 6.48 (*a*) ZVT PWM switching cell. (*b*) ZCT PWM switching cell.

ZVT PWM switching cell, and Fig. 6.48(*b*) shows the ZCT PWM switching cell. L is a resonant inductor, C is a resonant capacitor, S_1, is an auxiliary switch, and D_1, is an auxiliary diode.

6.7.2 The Boost ZVT PWM Converter

In this section, we consider the boost ZVT PWM, shown in Fig. 6.49, by placing the ZVT PWM switching cell shown in Fig. 6.48(*a*) into the conventional boost converter.

The switching cycle is divided into six modes, as shown in Fig. 6.50. These modes are defined as follows.

Mode I **($t_0 \leq t < t_1$)** Mode I, shown in Fig. 6.50(*a*), starts at $t = t_0$, when the auxiliary switch S_1 is turned on. Since the main switch, S, and the auxiliary switch, S_1, were off prior to $t = t_0$, it is clear that the diode, D, must have been on for $t < t_0$ to carry the output current. Hence, we assume D is on and D_1 is off at $t = t_0$. So for $t > t_0$, S_1 is on. The diode current starts to decrease, and it reaches zero when the inductor current i_L increases and reaches the constant current source, I_{in}, at t_1. In this mode, the capacitor voltage, v_c, is equal to the output voltage, V_o, and also equal to the inductor voltage as given by

$$V_o = L\frac{di_L}{dt} = v_c(t)$$

From the equation, the inductor current, i_L, is given by

$$i_L(t) = \frac{V_o}{L}(t - t_0)$$

This equation assumes zero initial condition for i_L.

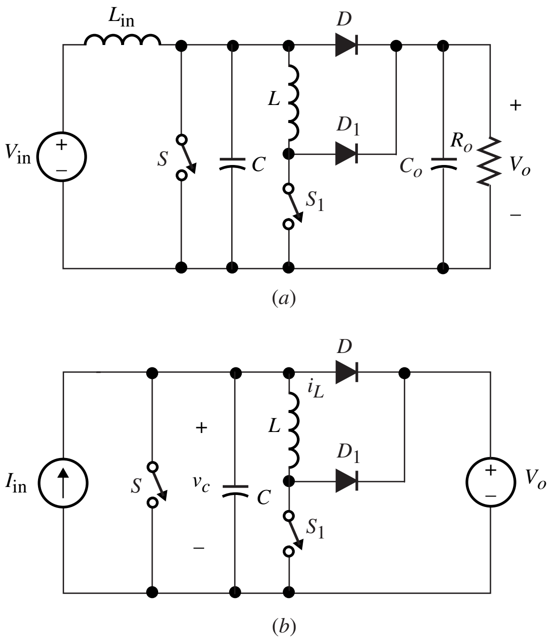


Figure 6.49 (*a*) Boost ZVT PWM. (*b*) Simplified equivalent circuit.

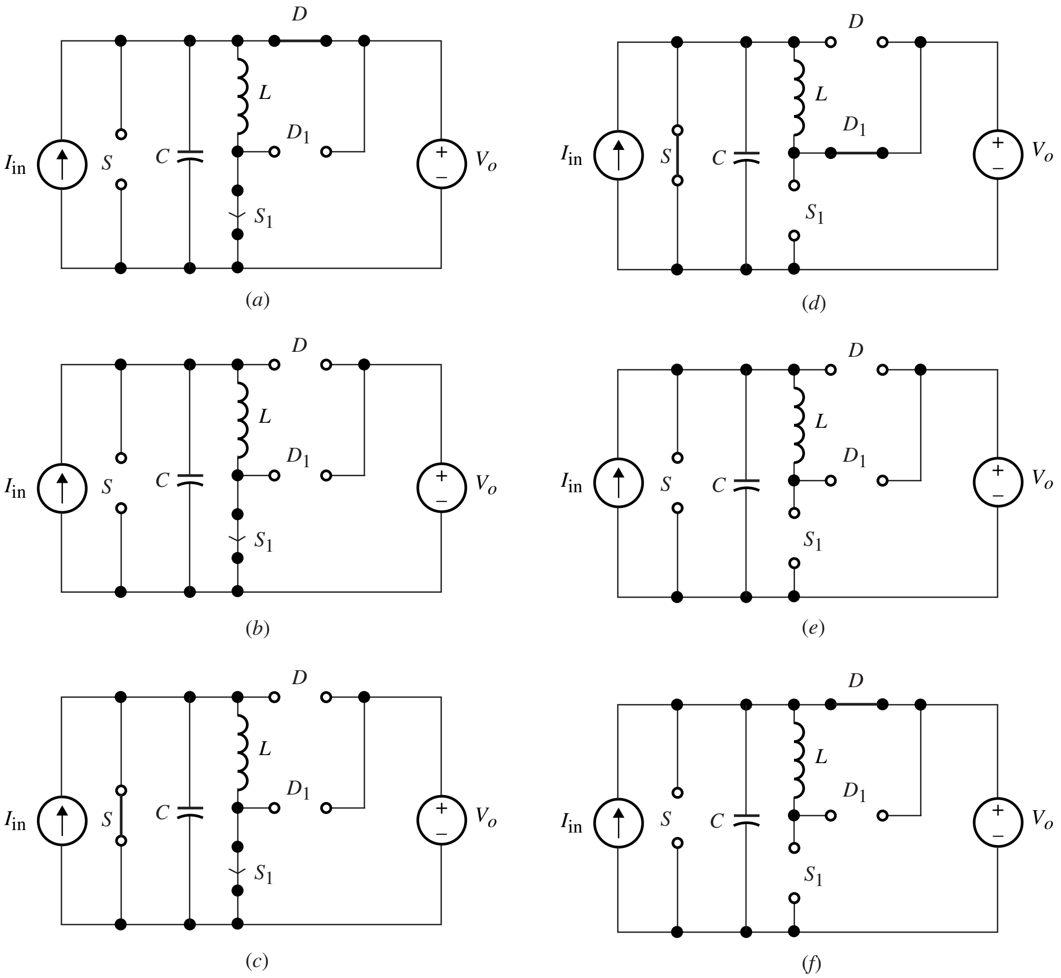


Figure 6.50 Equivalent circuits for the six modes of operation: (*a*) mode I, (*b*) mode II, (*c*) mode III, (*d*) mode IV, (*e*) mode V, and (*f*) mode VI.

As long as the inductor current is less than I_{in}, the diode will continue conducting and the capacitor voltage will remain at V_o. At time t_1, the inductor current becomes equal to I_{in}, D stops conducting, and the circuit enters mode II. From the above equation, we have

$$I_{in} = \frac{V_o}{L}(t - t_0)$$

The time interval is given by

$$(t_1 - t_0) = \frac{LI_{in}}{V_o}$$

This is the inductor current charging state.

Mode II **($t_1 \leq t < t_2$)** Mode II starts at t_1, when D is off. The circuit is shown in Fig. 6.50(b) and results in a resonant stage between L and C. During the time between

t_1 and t_2, the main switch, S, remains off and S_1 is still on, but both diodes are off. The initial capacitor voltage is still V_o, but the initial i_L has changed to I_{in}. The first-order differential equations that represent this mode are given by

$$C\frac{dv_c}{dt} = I_{\text{in}} - i_L(t)$$

$$L\frac{di_L}{dt} = v_c(t)$$

Equation (6.109) is obtained from the above two equations:

$$\frac{d^2 i_L}{dt^2} - \frac{1}{LC} i_L(t) = \frac{1}{LC} I_{\text{in}} \tag{6.109}$$

The solution for i_L and v_c is given by

$$i_L(t) = \frac{V_o}{Z}\sin\omega_o(t - t_1) + I_{\text{in}}$$

$$v_c(t) = V_o(2 - \cos\omega_o(t - t_1))$$

The time interval between t_1 and t_2 is given by

$$(t_2 - t_1) = \frac{1}{\omega_o}\cos^{-1}2$$

The diode voltage starts to charge up due to the decreasing capacitor voltage:

$$v_D(t) = V_o - v_c(t)$$

Substituting for v_c, the diode voltage becomes

$$v_D(t) = V_o(\cos\omega_o(t - t_1) - 1)$$

Mode III **($t_2 \le t < t_3$)** Mode III starts when the capacitor discharges to zero. The body diode of S turns on to clamp v_c to zero. In this mode the main switch, S, remains off, the auxiliary switch, S_1, is still on, and both diodes are off. Now,

$$v_c(t) = 0$$

Mode IV **($t_3 \le t < t_4$)** Mode IV starts at $t = t_3$, when the main switch, S, is turned on and the auxiliary switch, S_1, is turned off. At t_3, the initial capacitor voltage is zero, and the inductor starts linearly discharging from $i_L(t_2)$ to zero during t_3 to t_4. The diode D remains off since its voltage is negative, but D_1 turns on at $t = t_3$ to carry the inductor current.

The input voltage is equal to the inductor voltage, and the output voltage is equal to the negative inductor voltage, v_L.

$$v_L(t) = L\frac{di_L}{dt} = -V_o$$

The inductor current for $t > t_2$ is given by

$$i_L(t) = -\frac{V_o}{L}(t - t_2) + I_L(t_2)$$

Mode V ($t_4 \le t < t_5$) In mode V, at $t = t_4$ both switches are off, and both diodes are also off. The inductor current is zero, and the input current is going only through the capacitor:

$$I_{in} = C\frac{dv_c}{dt}$$

The capacitor voltage can be expressed as

$$v_c(t) = \frac{1}{C}I_{in}(t - t_4)$$

The capacitor is charging up from zero and will reach the output voltage at $t = t_5$. The time interval is

$$(t_5 - t_4) = \frac{V_o C}{I_{in}}$$

The circuit enters mode VI at this point.

Mode VI ($t_5 \le t < t_6$) When the capacitor reaches the output voltage, D starts conducting, but in this mode, both switches are still off. The diode current will equal the input current immediately. At $t = t_5$, the capacitor voltage is equal to the output voltage until the auxiliary switch is turned on again; then the cycle will repeat from mode I. The waveforms for the six modes of operation are shown in Fig. 6.51.

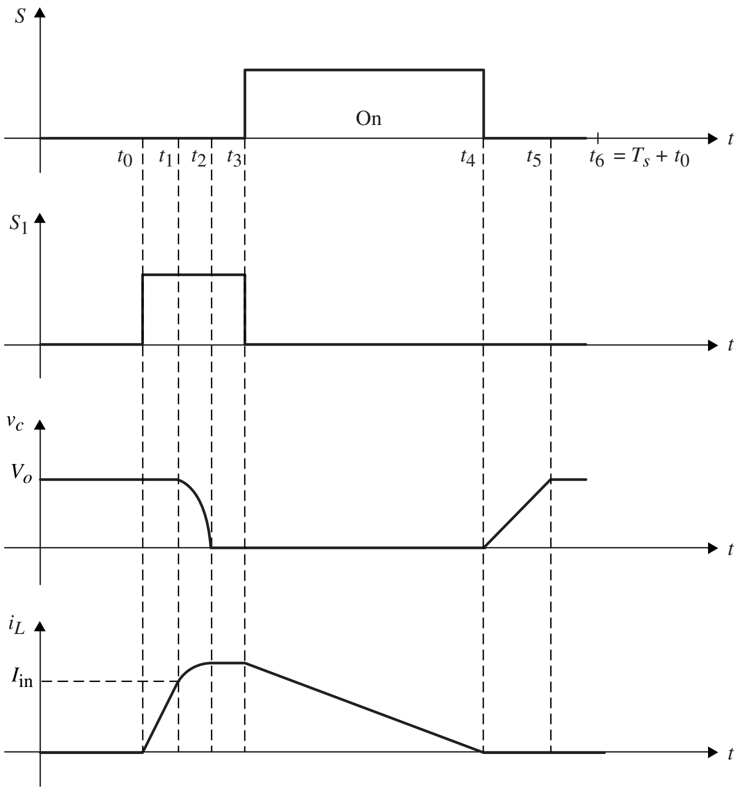


Figure 6.51 Steady-state waveforms for the ZVT boost converter of Fig. 6.49(*b*).

PROBLEMS

ZCS Quasi-Resonant Buck Converters

6.1 Consider the ZCS buck converter whose steady-state waveforms are shown in Fig. 6.10. Assume that the converter has the following parameters: $V_{in} = 25$ V, $I_o = 1$ A, $L = 3$ μH, $C = 0.02$ μF. Determine the time intervals at t_1, t_2, t_3.

6.2 Consider the ZCS buck converter shown in Fig. P6.2. Determine the output power.

6.3 Determine M and Q for a ZCS buck converter with an L-type switch that has the following converter components: $L = 15$ μH, $C = 60$ μF, $f_s = 100$ kHz, $V_{in} = 40$ V, and $I_o = 0.74$ A.

6.4 Consider the ZCS buck converter shown in Fig. P6.4(a) with $L = 20$ μH, $C = 5$ μF, $V_{in} = 50$ V, $V_o = 30$ V, $I_o = 20$ A. Assume that $L_o/R_o \gg 1/T$. The switching waveform of the transistor is shown in Fig. P6.4(b).

(a) Determine the minimum t_{on} in μs to achieve ZCS and determine the switching period T.

(b) Repeat part (a) by adding a diode across the transistor to allow bidirectional current flow, to produce a full-wave ZCS buck converter. Assume all values are the same as in part (a).

6.5 Determine the average output voltage for the ZCS buck converter shown in Fig. P6.5.

D6.6 Consider a ZCS buck converter with an L-type switch with $M = 0.35$ and $Q = 1$. Design for

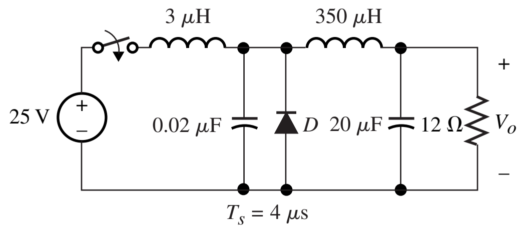


Figure P6.5

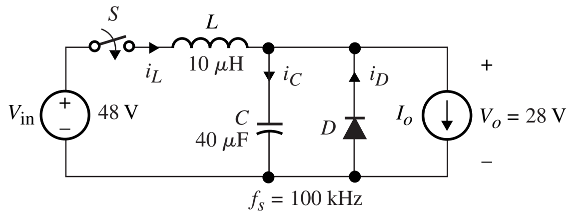


Figure P6.2

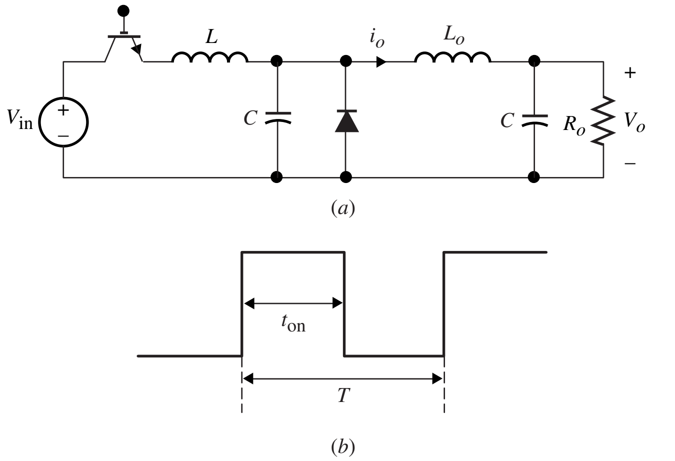


Figure P6.4

the resonant tank L and C for $V_{in} = 40$ V, $f_s = 250$ kHz, and $I_o = 0.7$ A.

D6.7 Consider the ZCS buck converter shown in Fig. P6.7 that has the following design parameters: $V_{in} = 25$ V, $V_o = 12$ V, $f_s = 250$ kHz, $I_o = 1$ A, and $f_o = 625$ kHz. Design for L and C and draw the steady-state waveforms for i_L, i, v_c, v_{sw}, v_D, i_D, i_o.

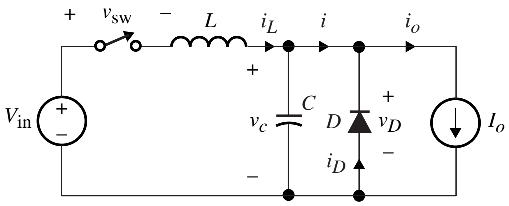


Figure P6.7

D6.8 Consider a ZCS buck converter with a unidirectional switch and the following specifications:

- Maximum load power 750 W
- Nominal output voltage 5 V
- Nominal input voltage 12 V
- Switching frequency 85 kHz
- Maximum peak resonant inductor current at twice the average load current

(a) Design for the resonant tank L and C.

(b) If the load current changes by ±50%, what are the new minimum and maximum switching frequencies required to maintain the output voltage at 5 V?

(c) Repeat part (*b*) for a variation of ±20% in the average input voltage.

D6.9 Given a ZCS buck converter with an L-type switch that has the following parameters: $V_{in} = 50$ V, $f_s = 100$ kHz, $I_o = 0.2$ A, $V_o = 49$ V, and $f_{ns} = 0.5$. Design for L, C, L_o, C_o, and R. Design for an output ripple current within 15% of its dc value and an output ripple voltage not to exceed 2%.

D6.10 Consider the ZCS buck converter shown in Fig. P6.10 with the following parameters: $L = 25$ μH, $C = 4.7$ μF, $V_{in} = 58$ V, and $I_o = 18$ A. Determine the output voltage and f_s to deliver an average output power equal to 580 W.

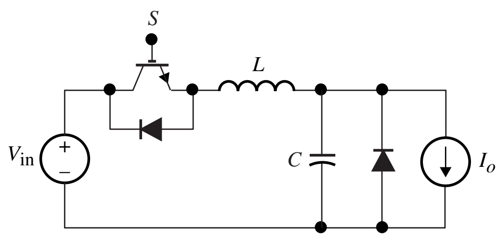


Figure P6.10

ZCS Quasi-Resonant Boost Converters

D6.11 Design a ZCS boost converter for the following parameters: $V_{in} = 20$ V, $V_o = 36$ V, $I_o = 0.7$ A, and $f_s = 250$ kHz. What is the range of f_s needed to keep V_o constant when I_o changes between 0.2 A and 1 A?

D6.12 Design a ZCS boost converter with an M-type switch with the following design parameters: $M = 2.73$ and $f_{ns} = 0.6$. It is desired to deliver a 0.5 A load current to $V_o = 68$ V.

D6.13 Design an M-type switch boost quasi-resonant converter with the following parameters: $M = 1.8$, $Q = 2.5$, $f_{ns} = 0.5$. The input voltage varies between 18 V and 26 V, and I_o varies between 0.5 A and 1 A. What is the range of f_s needed to keep the converter output voltage constant at $V_o = 38$ V?

ZCS Quasi-Resonant Buck-boost Converters

6.14 Consider the ZCS buck-boost converter with an L-type switch shown in Fig. P6.14. Find *M, Q* and f_s. What is the new f_s needed to keep V_o constant if the average load current changes to 1.8 A?

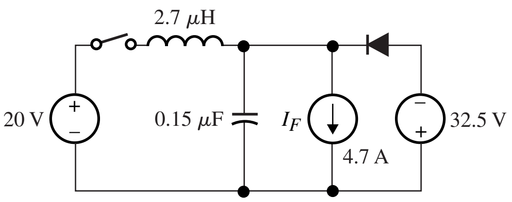


Figure P6.14

6.15 Determine the value of I_F in the ZCS buck-boost converter given in Fig. P6.15.

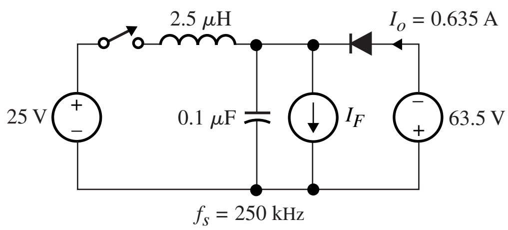


Figure P6.15

D6.16 Design a ZCS buck-boost converter with an L-type switch for the following specifications: $V_{in} = 40$ V, $V_o = 20$ V, $I_o = 4$ A, and $f_s = 250$ kHz.

D6.17 Design a ZCS buck-boost converter with L-type converter for the following specifications: $V_{in} =$ 30 V, $V_o = 20$ V @ $I_o = 2.8$ A, and $f_s = 50$ kHz.

D6.18 Design a ZCS buck-boost converter with an L-type switch that operates at 150 kHz and delivers 48 W to $V_o = 47$ V from a 10 V dc source.

D6.19 Figure P6.19 shows a ZCS buck-boost with *S* being unidirectional.

(a) Sketch i_L and v_c.

(b) Discuss the four modes of operation.

(c) Derive the expressions for i_L and v_c in terms of the circuit parameters.

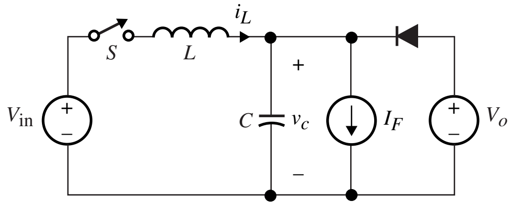


Figure P6.19

6.20 Repeat Problem 6.19 using a ZCS buck-boost L-type converter. Select any set of *M, Q,* f_{ns} you see fit for your design.

ZVS Quasi-Resonant Buck Converters

6.21 Derive the voltage gain expression for the M-type ZVS buck converter.

6.22 Consider the buck ZVS shown in Fig. P6.22 with an alternative way of implementing the L-type resonant switch. Assume L_o/L is very large.

(a) Discuss the modes of operation over one switching cycle.

(b) Draw the typical waveforms for i_L and v_c.

(c) Derive the expression for V_o/V_{in}.

6.23 Consider the resonant capacitor voltage and inductor current waveforms shown in Fig. P6.23 for a ZVS buck. Determine *L, C,* t_1, t_3, V_o, *Q*, f_{ns}, *M,* and $i_L(t_2)$.

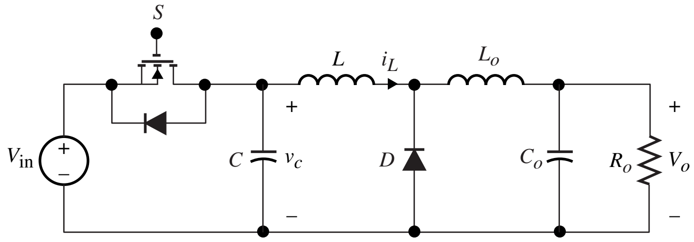


Figure P6.22

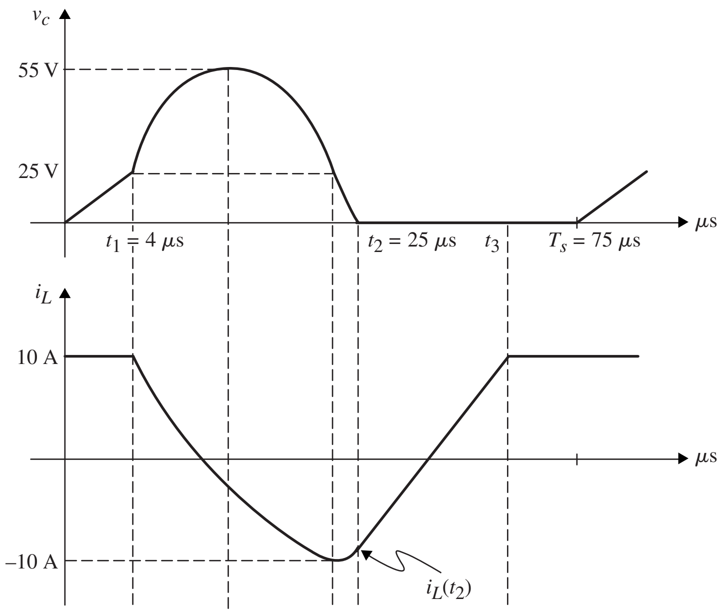


Figure P6.23

ZVS Quasi-Resonant Boost Converters

6.24 Consider the simplified ZVS boost quasi-resonant converter shown in Fig. P6.24. Assume S is a bidirectional switch.

(a) Sketch the waveforms for v_c and i_L.

(b) Discuss the four modes of operation.

(c) Derive the following expression for the voltage gain:

$$M = \frac{V_o}{V_{\text{in}}} = \frac{1}{\frac{f_{ns}}{2\pi}\left(\alpha - \frac{Q}{2M} + \frac{M}{Q}(1-\cos\alpha)\right)}$$

where

$$\sin\alpha = \frac{-Q}{M}, f_{ns} = \frac{f_s}{f_o}, Q = \frac{R_o}{Z_o}$$

$$Z_o = \sqrt{L/C}\,,\ f_o = \frac{1}{2\pi\sqrt{LC}}$$

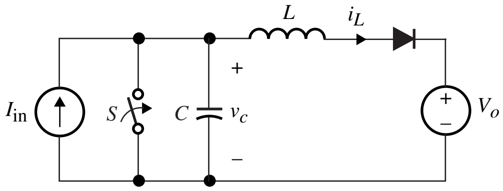


Figure P6.24

D6.25 Design a ZVS boost converter with the following specifications:

- Maximum load power 750 W
- Nominal output voltage 12 V
- Nominal input voltage 5 V
- Switching frequency 100 kHz
- Maximum peak capacitor voltage at 1.5 times the average input voltage

6.26 Figure P6.26(a) shows an isolated boost converter with switches that operates at 50% duty cycle. If it is assumed that L_{in} and C_o are large, then the input current, I_{in}, and the output voltage, V_o, may be assumed constant as shown in Fig. P6.26(b). The switching waveforms for S_1 and S_2 are shown in Fig. P6.26(c). It can be shown that there are four modes of operation in the steady state with S_2 operating at ZVS.

(a) Discuss the modes of operation.

(b) Show that $I_o/I_{\text{in}} = n(1-D)$, where I_{in} and I_o are the average input and output currents.

ZVS Quasi-Resonant Buck-Boost Converters

6.27 Determine the output voltage for the ZVS converter shown in Fig. P6.27 and sketch the waveforms for i_L, v_c, and i_D.

D6.28 Design a buck-boost ZVS converter with the following steady-state operating points: $M = 0.55, f_{ns} = 0.3, Q = 0.1$. Assume $V_{\text{in}} = 30$ V, $T_s = 10\ \mu$s, and the output current is at least 1 A.

General Soft-switching Converters

4.28 Figure P6.29 shows a zero-current transition (ZCT) buck converter.

(a) Discuss the modes of operation and show the typical waveforms for all the currents and voltages. Assume I_o is constant.

(b) Give the expression for the circulating energy in the resonant tank.

(c) What are the major features of this converter?

6.30 Consider the circuit given in Fig. P6.30(a) that operates in the ZVS mode with S_1 and S_2 operating alternatively as shown in Fig. P6.30(b). Discuss the operation of the circuit and sketch the waveforms for i_{sw}, i_{LK}, v_{c1}, v_{c2}, i_{Lm}, and v_s.

6.31 Figure P6.31(a) shows a ZVS soft-switching capacitor voltage-clamped converter. Assume $C_o \gg C_1$ and $C_o \gg C_2$. It can be shown that in the steady state there are four modes of operation. Assume S_1 and S_2 are switched as shown in Fig. 6.31(b). Sketch the waveforms for i_L, v_{c1}, v_{c2}, and i_{D2}, and discuss the modes of operation.

6.32 Figure P6.32 shows a technique that is based on the concept of clamped-current (CC) soft switching. This converter is known as the ZCS quasi-squarewave resonant converter. It is assumed that L_g is large enough so its current may be represented as a current source. S and D_s form a unidirectional switch. This topology offers several advantages. Discuss these advantages.

(a) Sketch the steady-state waveforms for i_L, i_{sw}, i_D, v_c, and v_o. Assume $I_F < I_o$ and the first mode starts at $t = t_0$ when S is on and D is on initially. It can be shown that there are four modes of operation.

(b) Show that during the resonant mode (mode 3), the resonant inductor current and resonant capacitor voltage equations are given by

$$i_L(t) = I_F(\cos\omega_o(t-t_2)+1)$$

$$v_c(t) = V_{in} + Z_o I_F \sin\omega_o(t-t_2)$$

where $Z_o = \sqrt{L/C}$ and $t = t_2$ is the start of mode 3.

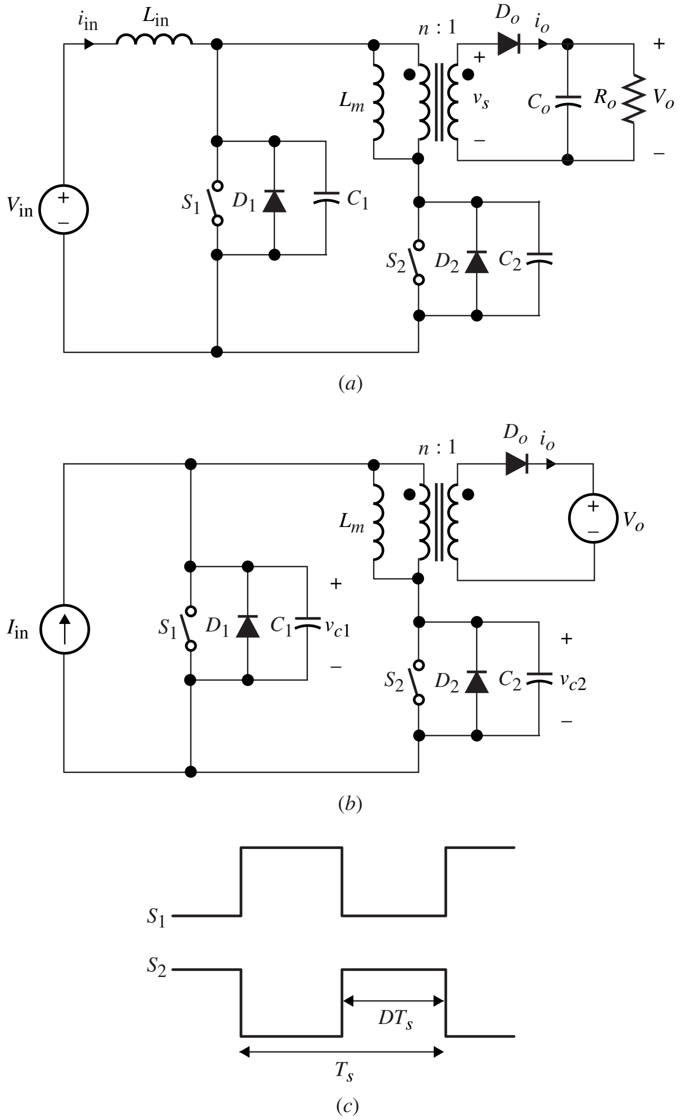


Figure P6.26

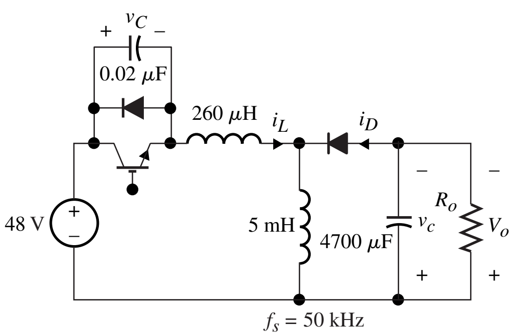


Figure P6.27

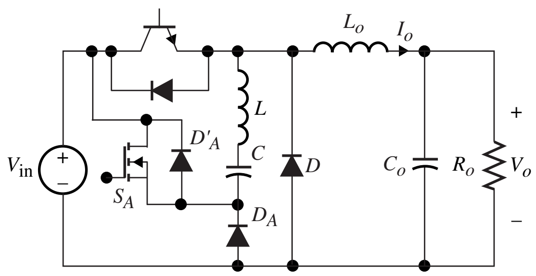


Figure P6.29

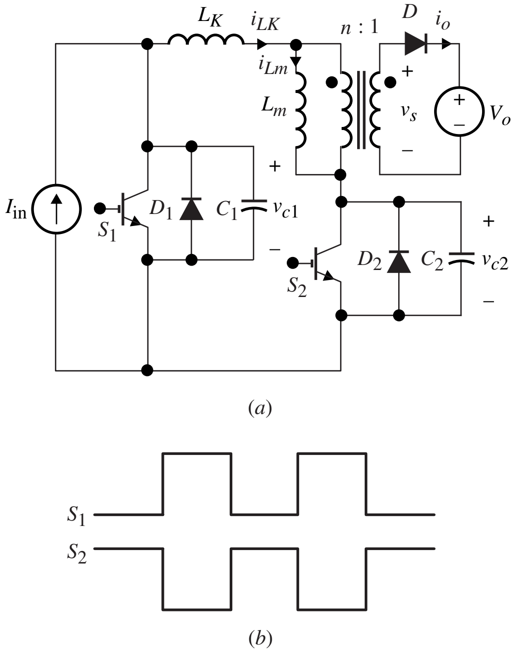


Figure P6.30

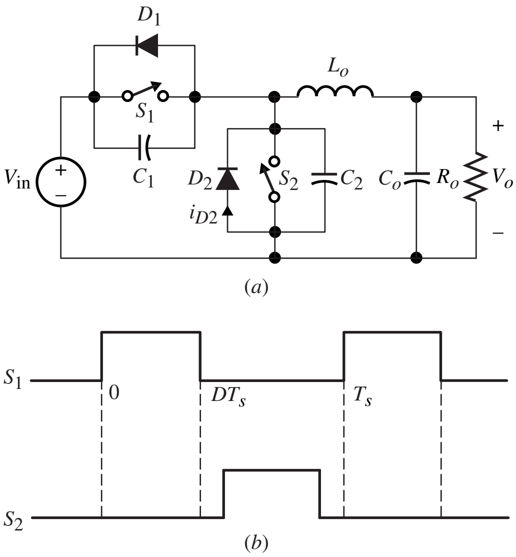


Figure P6.31

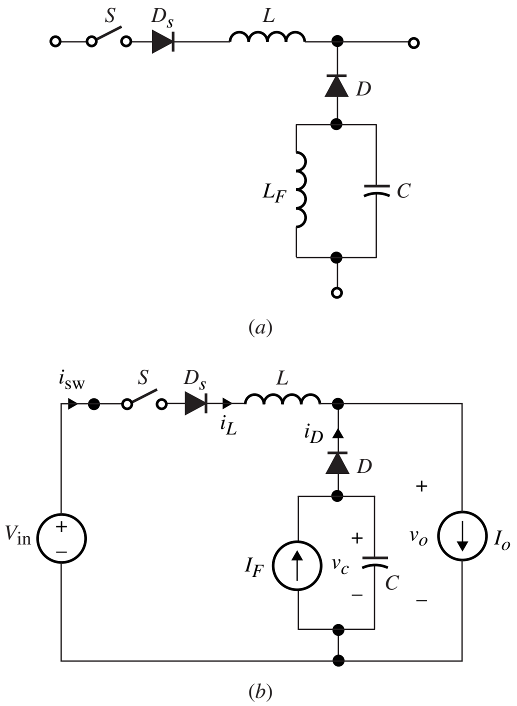


Figure P6.32

6.33 Figure P6.33 shows a soft-switching converter cell with the diode replaced by a switch S_D. By allowing the rectifier to become a controllable switch, the converter voltage gain can be controlled by the pulse width of S. This soft-switching family is known as PWM QUASI-SQUAREWAVE (PWM QSW) converters. Unlike the quasi-resonant converters, which are frequency-controlled converters, the QSW converters are PWM-controlled at constant frequency. Assume S and S_D are bidirectional switches. Discuss the four modes of operation.

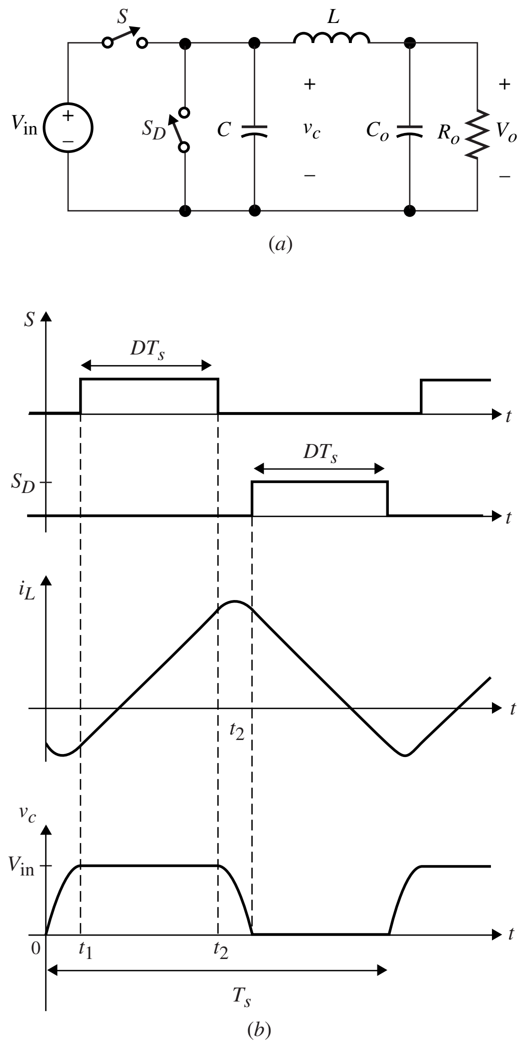


Figure P6.33