

Principles of Power Electronics

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Introduction

IN this chapter we describe power electronics and present a brief introduction to semiconductor switching devices and magnetic components. An introduction to these circuit elements is necessary because we utilize them in Part I, although we do not discuss them in detail until Part III. We also introduce nomenclature that we use throughout the book.

1.1 POWER ELECTRONIC CIRCUITS

The dominant application of electronics today is to process information. The computer industry is the biggest user of semiconductor devices, and consumer electronics, including cameras, is second. While all these applications require power (from a wall plug or a battery), their primary function is to process information; to take the digital optical signal produced by a compact disk and transform it into an analog audio signal, for instance. Power electronic circuits are principally concerned with processing energy. They convert electrical energy from the form supplied by a source to the form required by a load. For example, the part of a computer that takes the ac mains voltage and changes it to the 5-V dc required by the logic chips is a power electronic circuit (often abbreviated as *power circuit*). In many applications the conversion process concludes with mechanical motion. In these cases the power circuit converts electric energy to the form required by the electromechanical transducer, such as a dc motor.

Efficiency is an important concern in any energy processing system, for the difference between the energy into the system and the energy out is usually converted to heat. Although the cost of energy is sometimes a consideration, the most unpleasant consequence of generating heat is that it must be removed from the system. This consideration alone dictates the size of power electronic apparatus. Therefore a power circuit must be designed to operate as efficiently as possible. The efficiency of very large systems exceeds 99%. High efficiency is achieved by using

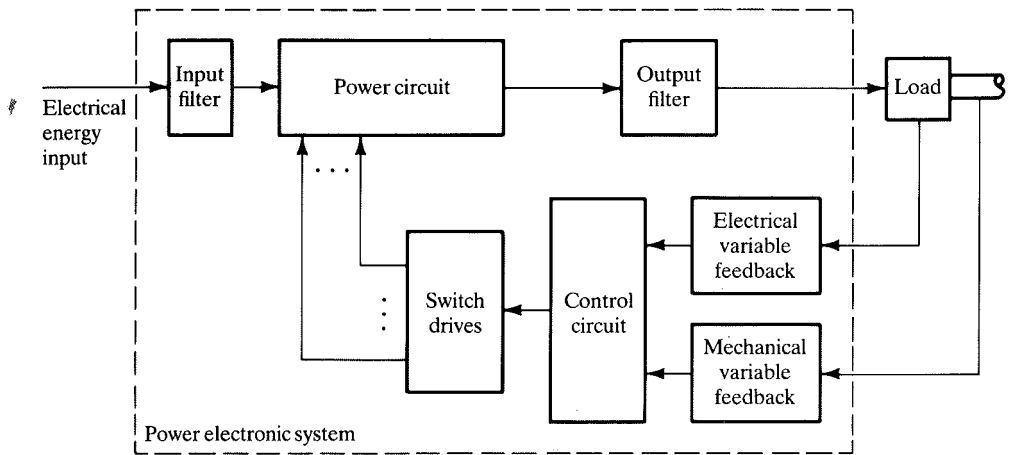


Figure 1.1 A block diagram of a typical power electronic system.

the power semiconductors as switches (where their voltage is nearly zero when they are on, and their current is nearly zero when they are off) to minimize their dissipation.[†] The only other components in the basic power circuit are inductors and capacitors, so the ideal power circuit is lossless.

A power electronic system consists of much more than a power circuit. The block diagram of a typical system is shown in Fig. 1.1. Switching creates waveforms with harmonics that may be undesirable because they interfere with proper operation of the load or other equipment, so filters are often employed at the inputs and outputs of the power circuit. The system load, which may be electrical or electromechanical, is controlled via the feedback of electrical and/or electromechanical variables to a control circuit. This control circuit processes the feedback signals and drives the switches in the power circuit according to the demands of these signals. The system also includes mechanical elements, such as heat sinks and structures to support the physically large components of the power circuit.

1.2 POWER SEMICONDUCTOR SWITCHES

The basic semiconductor devices used as switches in power electronic circuits are the bipolar and Schottky diodes, the bipolar junction transistor (BJT), the metal-oxide-semiconductor field-effect transistor (MOSFET), and a class of latching bipolar devices known as thyristors, the most common of which is the silicon controlled rectifier (SCR). Their circuit symbols and operating regions in the v - i plane are shown in Fig. 1.2. We discuss these and other hybrid devices in detail in Part III.

[†]Exceptions, such as linear voltage regulators, are so few that we do not consider them explicitly in this book.

Form and Function: An Overview

POWER electronic circuits change the character of electrical energy: from dc to ac, from one voltage level to another, or in some other way. We refer to such circuits generically as *converters*, *static converters* (because they contain no moving parts), *power processors*, or *power conditioners*. The part of the system that actually manipulates the flow of energy is the *power circuit*. It is the frame for the rest of the system's components, such as the control circuit or the thermal management parts.

The power circuit has a basic topology to which we add other circuit elements that perform ancillary functions, such as protection against transient overvoltages and filtering to eliminate electromagnetic interference. Although these other elements are important, they do not affect the function of the power circuit. Their purpose is to modify certain aspects of the power circuit's behavior, such as the rates of rise of currents or voltages. The study of a power circuit with all these additional elements can quickly become dominated by particulars rather than fundamentals. Therefore we concern ourselves only with the basic forms of power circuits in Part I of this book. We describe methods of building on these basic structures to transform them into practical power circuits, and eventually into systems, in Parts II, III, and IV.

In Part I we show how a desired conversion function influences the form, or topology, of the power circuit. We also use these forms to illustrate the analytic tools and ways of thinking that you should apply when studying a power circuit. In most cases we keep both the function and the form simple. Where we present a more advanced topology, our goal is to show the connection between it and its simpler form—and the benefits gained from the added complexity.

2.1 THE FUNCTIONS OF A POWER CIRCUIT

Before we can specify the form of a power circuit, we must define its function. In general, its function is to alter the characteristics of electrical energy provided by

one external system to those required by another. For instance, the power supply for a computer must convert the sinusoidal mains energy (60 Hz, 110 V rms in the United States) to a 5 V dc waveform. Another example is a power circuit for driving a variable speed ac motor, which might draw power from a battery and deliver a sinusoidal current waveform to the motor.

The types of functions that a power circuit can perform are limited only by the characteristics of electric power that are to be altered. As already mentioned, the transformations from ac to dc and from dc to ac are two possible functions. Interfacing systems with waveforms that are similarly shaped, but have different amplitudes, is another. If both waveforms are constant in time, we call the power circuit a *dc/dc converter*, and if both are alternating, we call the circuit an *ac/ac converter*. In the latter case, we might want to change the frequency or phase as well as the amplitude.

It is important that you not think of a power circuit's function as fixed for all time. The value of such a circuit is not just its ability to alter the form of electric energy, but also its ability to do so in response to a control signal. For instance, we can make the output of a computer power supply remain at 5 V even though the amplitude of the utility waveform changes by more than $\pm 20\%$. In some applications, such as light dimmers, the entire function of the power circuit is to provide this controllability.

A power circuit provides an interface between two other systems external to it and therefore imposes relationships between the voltage and current waveforms at one port and those at the other. Exactly what waveforms exist at these ports depends not only on these relationships, but also on how the two external systems respond to being related in this manner. Thus it is important always to describe the operation of a power circuit in the context of the external systems to which it is connected.

In certain situations one external system, such as a voltage source, will dictate a waveform at one port independent of the power circuit or the other external system. In these cases, we treat the waveform as an input to be processed by the power circuit to create an output. The external system at the output then defines the covariable (the current in this case) at its port, and we work backward through the power circuit to determine the shape of the covariable's waveform at the input. Because of their simplicity, we often use these cases when presenting the initial topologies in the following chapters.

The direction of energy flow usually determines whether ports serve as input or output. However, many power circuits are capable of processing bidirectional power, and identifying their ports as input or output leads to ambiguity. This ambiguity is aggravated by the fact that the types of semiconductor devices used to construct the circuit also constrain the direction of power. Thus we could construct two identical topologies to process power in opposite directions. This is an important issue, which we address frequently, for it emphasizes that two visually distinct circuits can behave similarly.

Figure 2.5(b) shows the waveforms v_a and v_{ac} when the filter is tuned to the fundamental of v_a . That is, the switching frequency ω_s is equal to the resonant frequency $\omega_o = 1/\sqrt{LC}$. If the Q of the series RLC circuit is high, the filter provides good selectivity and v_{ac} is nearly sinusoidal. At switching frequencies other than $\omega_s = \omega_o$, we can determine the amplitude of the output from the magnitude of the admittance $Y(j\omega_s)$:

$$\begin{aligned} |V_{ac}| &= RY(j\omega_s)V_{a_1} \\ &= R \left| \frac{-j\omega_s C}{\omega_s^2 LC - j\omega_s RC - 1} \right| \left(\frac{4}{\pi} V_{dc} \right) \end{aligned} \quad (2.3)$$

where V_{a_1} denotes the amplitude of the fundamental (ω_s) component of v_a .

The magnitude of $Y(j\omega)$ is shown in Fig. 2.6. Switching at a frequency higher than ω_o allows the filter to still do a good job of removing harmonics from v_a . The rapid attenuation provided by the filter reduces the almost sinusoidal output voltage. Thus by varying the switching frequency, we can control the power delivered to the load resistor.

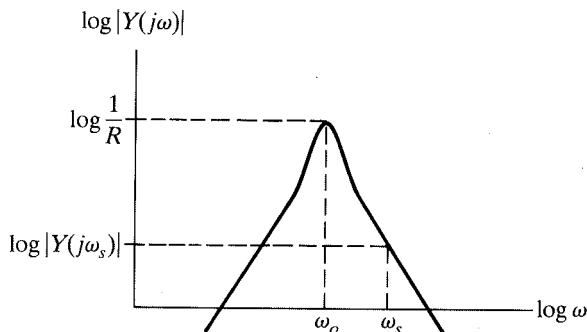


Figure 2.6 Admittance as a function of frequency for the RLC circuit on the ac side of Fig. 2.5(a).

The topology of Fig. 2.5(a) is known as a *series resonant converter*. It is one member of a family of ac/dc converters called *resonant converters*. Designers use them principally to obtain clean, high-frequency ac waveforms, often for induction heating applications. Note that the technique of power control that we just discussed results in a varying output frequency. Therefore any application of this type of control must not involve a precise output frequency.

2.3 DC/DC CONVERTERS

Used extensively in power supplies for electronic equipment, dc/dc converters control the flow of energy between two dc systems. The dc/dc converter takes the dc output of the ac/dc converter and transforms it to the different dc voltages required by the electronics—5 V and ± 15 V, for example. These converters are also used in battery powered equipment and to control the speed of dc motors in many traction applications, such as battery powered forklifts or trains operating from a dc third

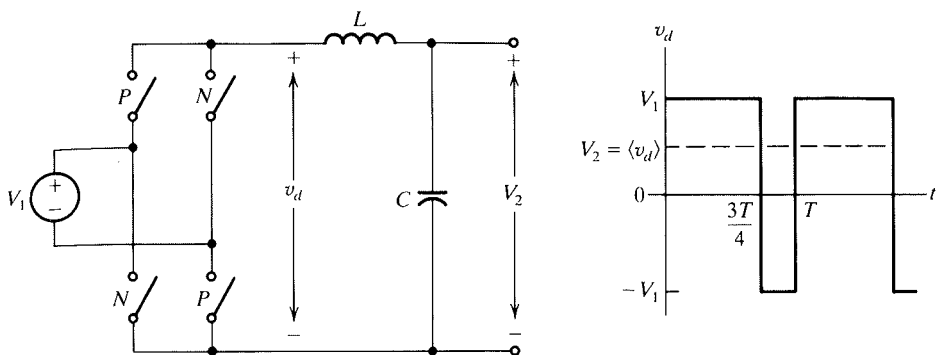


Figure 2.7 Basic dc/dc converter topology. The conversion ratio V_2/V_1 can vary between ± 1 .

rail or catenary. In relatively high-power applications, such as traction, the dc/dc converter is known as a *chopper*.

2.3.1 Basic Topology

The basic dc/dc converter topology is shown in Fig. 2.7. The striking feature of this circuit is its similarity to the ac/dc converter topology shown in Fig. 2.1(b). In fact, the only difference is the control of switches. In the dc/dc converter, the switches are controlled to produce a voltage v_d that contains a nonzero dc component. This component is then extracted by the low-pass LC filter to produce the output dc voltage V_2 .

As before, we alternately operate the P and N switches. Instead of leaving each set closed for exactly half the cycle as we did in the ac/dc converter, here we control them to have asymmetric on-times so that v_d contains a dc component. For example, if we leave the P switches on for three quarters of the switching period T , the voltage v_d shown in Fig. 2.7 results. It has an average value of $V_1/2$. In this circuit we can obtain an average voltage between V_1 and $-V_1$ by adjusting the relative conduction periods of the two sets of switches. The fraction of its switching period during which a switch is on is known as the *duty ratio* D of the switch. Here the duty ratio of the P switches is 0.75, and that of the N switches is 0.25.

EXAMPLE 2.3

A Simplified dc/dc Converter Topology

We can simplify the topology of Fig. 2.7 by requiring an output voltage of only one polarity; for example, $0 < V_2 < V_1$. The circuit of Fig. 2.8, which contains only two switches, is sufficient to do the job. As we turn these switches on and off sequentially, the voltage waveform they create steps between V_1 and 0.

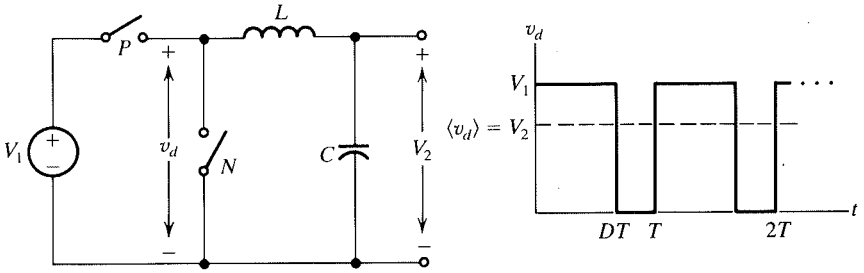


Figure 2.8 A simplified dc/dc converter topology limited to producing an output voltage $0 < V_2 < V_1$.

Assume that we leave switch P on for time DT , where T is the switching period and D is the duty ratio of the P switch. We then turn switch P off and switch N on for the rest of the period, that is, for a time $(1 - D)T$. The output voltage V_2 , the average value of v_d , is then DV_1 . We can alter this average value by changing the duty ratio D , but the lower limit of the output voltage is zero.

What did we gain from this simplification? One benefit is the reduction in the number of switches required. Another, which can be shown through a harmonic analysis, is that the ac components of v_d in this circuit are smaller than those in v_d for the bridge circuit of Fig. 2.7. We can therefore use smaller filter elements to achieve the same level of ripple in the waveforms presented to the external system.

Because the frequency of the desired output (dc) of the converter of Fig. 2.7 is much less than the switching frequency ($1/T$), it is a member of a class of converters known as *high-frequency switching converters*. We discuss high-frequency dc/dc converters in Chapter 6. As you will see in Chapter 8, another member of this class is the high-frequency ac/dc converter.

We can also achieve the transformation from dc to dc in a way different from that which we have been discussing. Instead we can use a dc/ac converter to create an ac waveform from the dc at one port and then an ac/dc converter to transform this ac waveform back to dc at the other port. At first glance this approach appears to be wasteful because it requires so many switches and filter elements. It does, however, give us a point in the circuit where the waveforms are ac. As shown in Fig. 2.9, we can install a transformer at this point to give electrical isolation and to

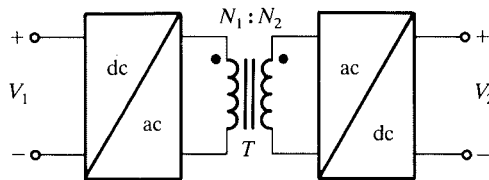


Figure 2.9 A dc/dc converter consisting of two cascaded ac/dc converters. This configuration permits the use of an isolating transformer T , as shown.

make use of its turns ratio when the difference between the two external voltages is large. We discuss this type of converter, called an *isolated high-frequency converter*, in Chapter 7.

2.4 AC/AC CONVERTERS

An ac/ac converter converts an ac waveform of one amplitude, phase, and/or frequency to another ac waveform with different parameters. As with the dc/dc converter, several different topologies can be used, depending on system requirements.

2.4.1 Topologies

We cannot identify a basic topology for the ac/ac converter, as we did for the ac/dc and dc/dc conversion functions. One of three distinct topological approaches can be used, depending on system requirements. The first and simplest can be used to change the amplitude parameters of an ac waveform (the rms value, for instance). It is known as an *ac controller* and functions by simply taking symmetric “bites” out of the input waveform. The second can be utilized if the output frequency is much lower than the input source frequency. This topology is called a *cycloconverter*, and it approximates the desired output waveform by synthesizing it from pieces of the input waveform. The third approach consists of two ac/dc converters with their dc ports connected. The result is known as a *dc link converter*.

The ac Controller Figure 2.10 illustrates the operation of the ac controller as a light dimmer. The switch is opened at every zero-crossing and reclosed at some later point in the half-cycle. This circuit does not control the fundamental frequency of the output waveform. However, it can control the rms value of the output or the amplitude of its fundamental component by varying the turn-on point, α , of the switch. To keep the harmonic components of the waveform created by the switch

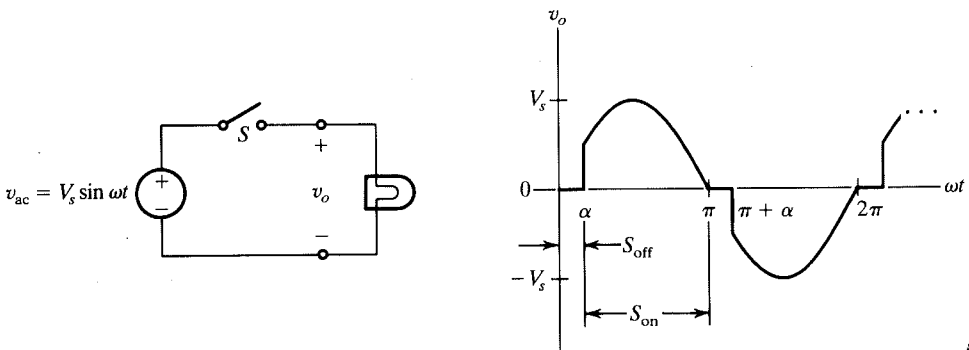


Figure 2.10 The ac controller utilized as an incandescent light dimmer.

Isolated High-Frequency dc/dc Converters

WE add transformers to the topology of a high-frequency converter for two reasons: (1) to provide electrical isolation between the two external systems; and (2) to reduce the component stresses that result when the input/output conversion ratio is far from unity. (We showed the relationship between switch stress factor and the conversion ratio in Fig. 6.23.) There are many ways in which we can include the transformer in the topology of a dc/dc converter. We present and discuss some of them in this chapter. We believe that the overview of transformers presented in Chapter 1 is sufficient background for understanding the material in this chapter. However, if you are still uncomfortable when confronted with a transformer, you should read Section 20.4 before continuing with this chapter.

Recall that a transformer winding cannot have a dc voltage across it, because the magnetizing inductor is a short circuit at zero frequency. Therefore we need to create, from the dc voltages of the external systems, an ac voltage with no average value. Thus we usually—but not necessarily—have the switches that produce this ac voltage also control the input/output conversion ratio.

There are essentially two high-frequency transformer topologies. The first, referred to as a *forward converter*, is based on the direct converter. The second, referred to as a *flyback converter*, is based on the indirect converter. In the forward converter, the algebraic sum of the instantaneous power over all windings is zero. That is, the transformer is not required to store significant energy. Although some energy is stored in the transformer's magnetizing inductance, we minimize this energy by making the inductance large. In the flyback converter, however, the transformer is required to store energy. During one part of the switching cycle, the primary winding takes energy from the input system and stores it in the magnetizing inductance. During the second part of the cycle, a second winding removes this energy and delivers it to the load. These two basic isolated converter topologies are the subjects of this chapter.

Even though it may not be immediately apparent, every topology presented in this chapter is based on the canonical cell of Section 6.2. The transformer, of

course, adds additional elements to the circuit, but we can identify basic isolated converter operation during periods DT and $(1 - D)T$ with one of the canonical cell connections.

7.1 THE SINGLE-ENDED ISOLATED FORWARD CONVERTER

Figure 7.1(a) shows one version of an isolated down converter. The transformer is modeled as an ideal transformer with a shunt inductor L_μ representing its magnetizing inductance. (For now we are going to ignore the transformer's leakage inductance.) We call this converter a *single-ended* isolated forward converter—"single-ended" because power flows through the transformer for only one polarity of the primary voltage. The diode D_3 and the voltage source V_c provide a path for the magnetizing current i_μ when the transistor turns off.

When the transistor turns on, the transformer primary voltage v_P is equal to the input voltage V_1 . The secondary voltage v_S is related to v_P by the turns ratio, and its polarity is positive. Therefore the output rectifier diode D_1 is on, the freewheeling diode D_2 is reverse biased, and $v_d = V_1/N$. The output current I_2 is reflected to the primary circuit as $i_1 = I_2/N$. Note that if we ignore the magnetizing current, and if $N = 1$, operation of the single-ended isolated forward converter is indistinguishable from the nonisolated down converter at this point. Furthermore, a nonunity turns ratio simply provides a step up or step down of the voltage and current.

If we ignore the magnetizing inductance, the transformer's primary and secondary currents i_P and i_S are zero when the transistor is off. The freewheeling diode D_2 will therefore turn on to carry the output current and $v_d = 0$. This condition corresponds to that part of the cycle when the shunt switch S_{xz} is conducting in the nonisolated direct converter of Fig. 6.8(a).

A practical isolated single-ended forward converter has two features that complicate both its performance and its design. The first is the need to provide for the consequences of nonzero stored energy in the transformer magnetizing branch L_μ . The second is the fact that processing this magnetizing energy results in switch stresses that exceed those of the nonisolated forward converter. We now consider these two features in detail.

7.1.1 Magnetizing Current and Clamping

If the transformer magnetizing current i_μ is nonzero, we can no longer ignore it when the transistor Q turns off. We must provide the current with a path in order to prevent it from being discontinuous. Furthermore, while i_μ is flowing during the time Q is off, v_P must be negative so that $di_\mu/dt < 0$ and $i_\mu(T) = i_\mu(0)$. This is the periodic steady state condition, and it guarantees that $i_\mu(T)$ is not increasing

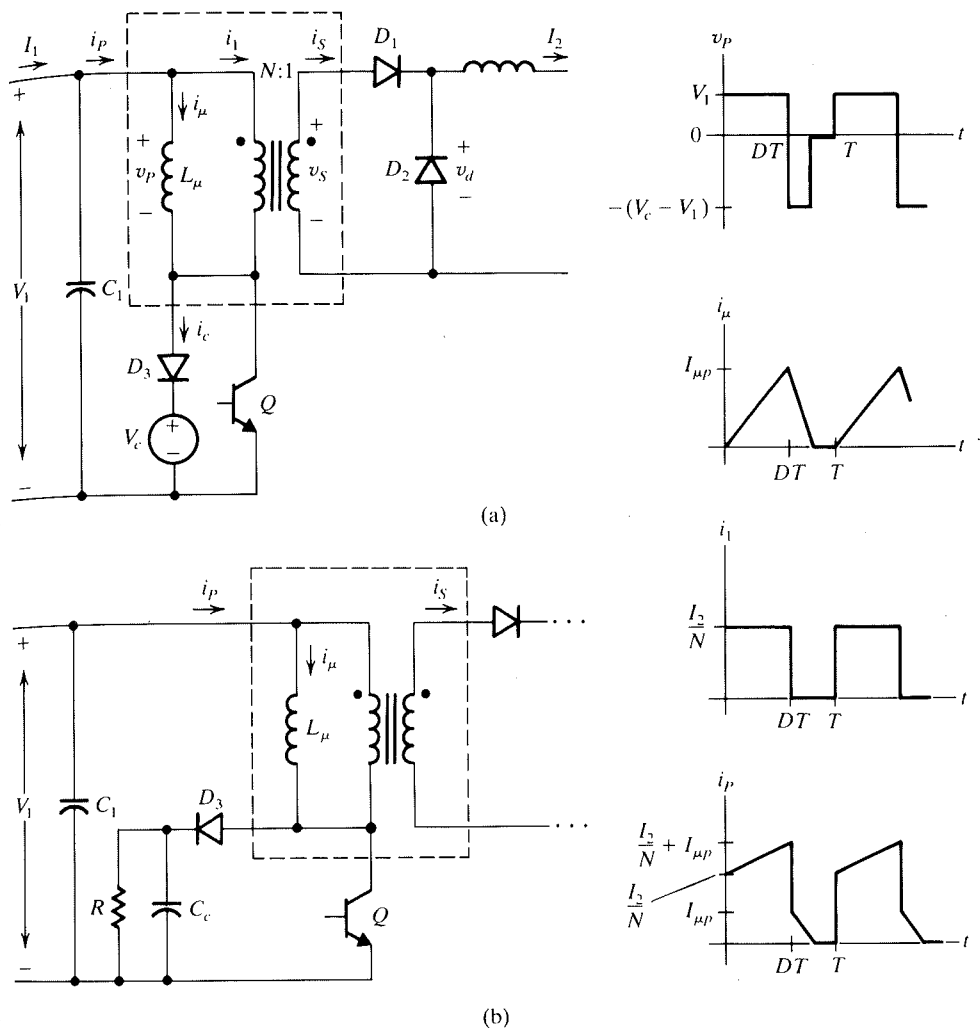


Figure 7.1 Single-ended isolated down converter with clamp: (a) clamp voltage represented by a voltage source V_c ; (b) clamp voltage implemented with a capacitor C_c and a discharge resistor R .

every cycle. Stated another way, the magnetic flux B in the core must return to its starting value at the end of every cycle. (This condition is called *resetting* the core.) Therefore we must provide a means for reversing the polarity of v_p when Q is off. We do so with the diode D_3 and the clamp voltage V_c .

You can best understand the operation of the clamp by first assuming that the

magnetizing current is zero when Q turns on. Because v_P is positive and constant at V_1 when Q is on, i_μ , which adds to the current carried by Q but not by D_1 , increases linearly. When Q turns off, I_2 immediately commutates to D_2 , and D_1 turns off. The ideal transformer currents i_S and i_1 step to zero at this time, but i_μ must be continuous. It cannot flow into the primary winding of the ideal transformer because the corresponding secondary current would have to flow the wrong way through D_1 . Instead, i_μ forces D_3 on, clamping the transistor collector at V_c and v_P at $V_1 - V_c$.

The magnetizing current i_μ must be zero by the start of the next cycle. If i_μ were not zero, it would continue to increase each cycle until the transformer saturated. If i_μ is to return to zero by the start of the next cycle, v_P must be negative when Q is off. The voltage V_c must therefore be greater than V_1 . How much greater depends on the relative lengths of time Q is on and off. Specifically, the time integral of v_P (generally referred to as the *primary volt-seconds*) across the transformer when Q is off must equal the negative of the primary volt-seconds when Q is on. That is, the net volt-seconds must be zero. For example, if Q is on for 50% of the cycle, V_c must equal $2V_1$. Note that this condition is the same as $\langle v_P \rangle = 0$.

If V_c were precisely equal to the value necessary to balance the positive and negative primary volt-seconds, the transformer flux at the end of each cycle would return to its value at the beginning of the cycle. Because of a transient or an imperfection in control, this starting value may not be zero. In such a case, the peak flux in the transformer will be higher than necessary, and we may still have to worry about saturation. Therefore, to guarantee that $i_\mu(0) = 0$, we usually make the clamp voltage slightly greater than the critical value necessary for balancing the volt-seconds. Diode D_3 will prevent i_μ from actually going negative, so the minimum flux level in the transformer is $B_{\min} = 0$. When D_3 turns off, v_P steps to zero, where it stays until the transistor turns on again.

The duty ratio is often a control variable, so we must choose V_c to guarantee that i_μ will be reset to zero in the worst case. For example, the larger the duty ratio in the converter of Fig. 7.1, the higher will be the positive volt-seconds seen by the transformer and the larger V_c must be. Therefore we must choose V_c on the basis of the expected maximum duty ratio. For duty ratios below this maximum value, i_μ will simply return to zero more quickly than necessary, and a greater amount of the transistor's off time will be spent with $v_P = 0$.

There are many different ways of implementing the clamp function, only some of which we discuss. In practice, we frequently use a breakover device—such as a Zener diode or a metal-oxide varistor—as a direct replacement for the diode and clamp voltage source of Fig. 7.1(a) if the energy involved is within the ratings of the breakover device. However, the principles involved are independent of the specifics of implementation. For this reason, we use easily understood clamp circuits to illustrate the important issues of the clamping function.

EXAMPLE 7.1**Choosing a Clamp Voltage**

Assume that the circuit of Fig. 7.1(a) has an input voltage $V_1 = 50$ V and that the transformer has a turns ratio $N = 5$. If the transistor's duty ratio is D , the output voltage will be

$$V_2 = D \left(\frac{V_1}{N} \right) = D(10) \text{ V} \quad (7.1)$$

If the requirement on V_2 is $0 < V_2 < 8$ V, we need a maximum duty ratio of 80%. We can find the minimum necessary value of V_c by equating the positive and negative volt-seconds on the primary:

$$V_1 D_{\max} T = (V_c - V_1)(1 - D_{\max})T \quad (7.2)$$

or

$$V_c = \frac{V_1}{1 - D_{\max}} = \frac{50}{0.2} = 250 \text{ V} \quad (7.3)$$

The clamp voltage should be somewhat greater than this minimum value. How much greater depends on how sure we are that the maximum duty ratio will not exceed 80% or how carefully we control transients that can cause flux offsets. Note that we must choose a transistor rated to withstand V_c , which is high relative to other voltages in the circuit.

Energy is absorbed by the clamping circuit every time it operates, that is, once a cycle. This energy is often difficult and uneconomical to recover from the source V_c , and hence it represents a loss. We can calculate the amount of this energy by integrating $V_c i_c$ over the time that D_3 is on. The voltage V_c is a constant, and the current i_c decreases linearly from its peak value of:

$$I_{\mu p} = \frac{V_1 D T}{L_\mu} \quad (7.4)$$

to zero over a time,

$$\Delta t = \frac{L_\mu I_{\mu p}}{V_c - V_1} \quad (7.5)$$

The energy flowing into the source V_c every cycle, E_c , therefore is

$$E_c = \int_0^{\Delta t} V_c i_c dt = V_c \frac{I_{\mu p}}{2} \Delta t = \frac{1}{2} L_\mu I_{\mu p}^2 \left(\frac{V_c}{V_c - V_1} \right) \quad (7.6)$$

which is larger than $L_\mu I_{\mu p}^2 / 2$, the energy stored in L_μ .

We usually develop the clamp voltage V_c by charging a relatively large capacitor. In order to prevent V_c from increasing every cycle, we must provide a

means to remove E_c . One simple solution is to place a resistor in parallel with the capacitor— R in Fig. 7.1(b). Because both the energy flowing into and out of the capacitor depend on V_c , we must choose exactly the right value of R to get the desired clamping voltage. To keep V_c constant, this resistor value has to change with a change in either V_1 or D . Therefore we generally use a linear network that senses and regulates V_c instead of a simple resistor of fixed value. If the application is a very high-power one, we might justifiably use an auxiliary dc/dc converter to recover E_c by transferring it to either the source or load.

EXAMPLE 7.2

A More Efficient Placement of the Clamp Discharge Resistor

By placing the clamp circuit regulator or discharge resistor in parallel with the clamp capacitor (assumed to be infinite), we dissipate an amount of energy given by (7.6). If we instead connect the resistor between the capacitor and the input voltage source, as shown in Fig. 7.2, part of the energy removed from the capacitor returns to the source. The ratio of the energy dissipated in the resistor every cycle, E_R , to that absorbed by the clamp, E_c , is equal to the ratio of the resistor voltage, $V_c - V_1$, to the clamp voltage, V_c . The energy dissipated is thus:

$$E_R = E_c \left(\frac{V_c - V_1}{V_c} \right) = \frac{1}{2} L_\mu I_{\mu p}^2 \left(\frac{V_c}{V_c - V_1} \right) \left(\frac{V_c - V_1}{V_c} \right) = \frac{1}{2} L_\mu I_{\mu p}^2 \quad (7.7)$$

Because the energy given by (7.7) is less than that given by (7.6), the resistor placement in Fig. 7.2 is more efficient than that in Fig. 7.1(b).

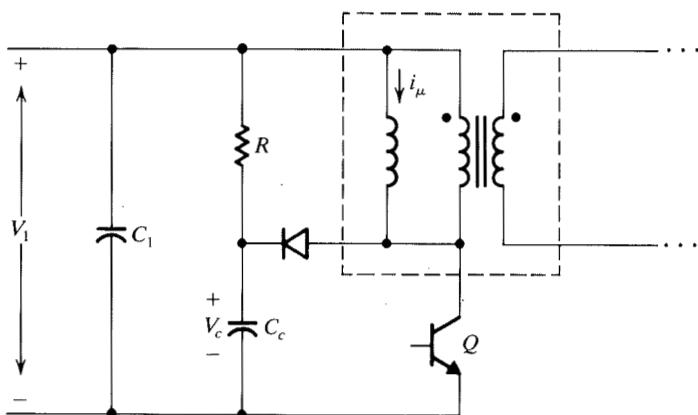


Figure 7.2 An alternative to the discharge resistor placement in Fig. 7.1(b).

7.1.2 A Transformer-Coupled Clamp

We show another way to provide the clamp function in Fig. 7.3, where a third winding of N_T turns has been added to the transformer. This third winding, called

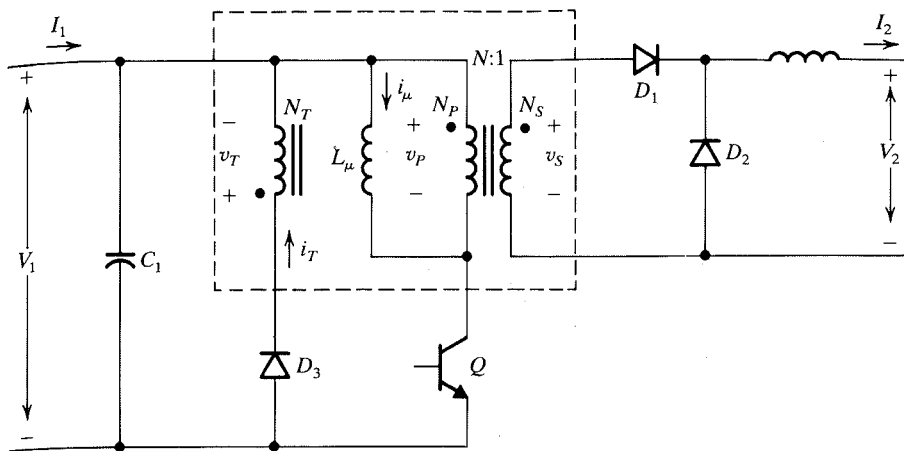


Figure 7.3 Clamp function provided by a tertiary transformer winding.

a *tertiary* or *clamp winding*, permits i_μ to circulate through the primary and clamp windings when Q turns off. The current i_T therefore is

$$i_T = \frac{N_P}{N_T} i_\mu$$

Note that when Q turns off, i_μ will flow out of the dot on N_P , causing i_T to flow into the dot on N_T and D_3 to turn on. Because $v_P = v_T(N_P/N_T)$ and $v_T = -V_1$ during the time when D_3 is on, L_μ discharges, as desired. However, i_T cannot go negative, and therefore we can still reduce the transformer flux only to zero. The advantage of this approach is that the magnetizing energy returns directly to the source, instead of to a separate clamp circuit.

We can adjust the effective clamp voltage (the voltage across L_μ when Q is off) by changing the ratio N_P/N_T . For a maximum duty ratio of 50%, N_P/N_T must have a minimum value of 1. For a maximum duty ratio of 75%, N_P/N_T must have a minimum value of 3. Note that this turns ratio affects the off-state voltage of the transistor. For the 1:1 ratio, the transistor's off-state voltage is $2V_1$, and for the 3:1 ratio it is $4V_1$.

A problem with the circuit of Fig. 7.3 is that any leakage inductance between the primary and tertiary windings keeps the magnetizing current from commutating immediately to the tertiary winding when the transistor turns off. An additional clamp, or a transient suppressing circuit called a *snubber*, must be placed across the transistor to keep its voltage from rising too much. (We discuss snubber circuits in detail in Chapter 24.) The energy associated with this leakage inductance typically is small compared to that of the magnetizing inductance, so the dissipation in this

clamp or snubber is not a problem unless the circuit is switching at a very high frequency.

Instead of connecting the clamp winding across V_1 , we could connect it across V_2 , transferring the magnetizing energy to the output rather than back to the input. The result is a slight improvement in the circuit's efficiency. Owing to safety isolation specifications that often call for a minimum spacing between source and load windings, however, this approach makes a tight coupling of the primary and tertiary windings difficult. The greater dissipation caused by the larger leakage inductance easily offsets the expected gain in efficiency.

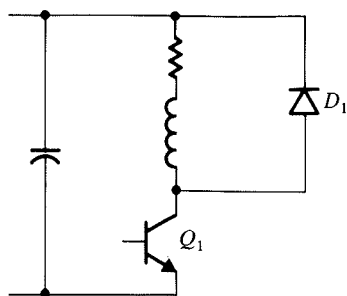
7.1.3 The Isolated Hybrid Bridge

A technique frequently used to maintain current continuity in inductive loads switched by transistors is to place a diode around the load, as shown in Fig. 7.4(a). The problem with this technique for our present purposes, however, is that the discharging voltage is only a diode drop plus the drop across the resistance of the winding. But the circuit of Fig. 7.4(a) suggests another possibility. When Q_1 is off, the bottom terminal of the load is connected to the positive rail through D_1 . Thus, if we could connect the top terminal of the load to the negative rail, we could apply $-V_1$ across the inductor to discharge it. We can do so by adding Q_2 and D_2 , as shown in Fig. 7.4(b). The transistors Q_1 and Q_2 turn on and off simultaneously. Note that again, because the diodes do not permit i_μ to reverse, the transformer core resets to a minimum flux value of only $B_{\min} = 0$. Because only two of the four bridge switches are controllable, this circuit is still only single-ended and is called a *single-ended isolated hybrid-bridge*.

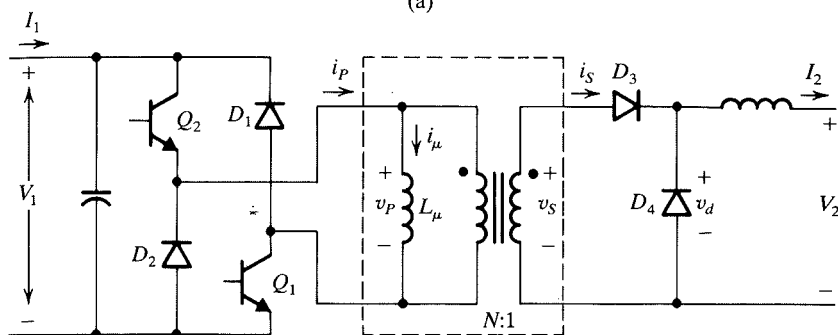
Although this converter has twice as many primary side switches as that of Fig. 7.3, they all have a smaller required voltage rating. And as the magnetizing current does not need to commutate from the primary winding to a clamp winding, we avoid the leakage inductance in the circuit of Fig. 7.3. We do need to drive two transistors whose emitters are at different potentials, however, and the circuit is limited to a maximum duty ratio of 50%. We can obtain this duty-ratio limit from the current waveforms shown in Fig. 7.4(c). These waveforms are based on the assumption that the transformer leakage inductance is zero. This is not true in a practical transformer: A commutation period results, during which both D_3 and D_4 are on, as we will discuss in Section 7.4.

7.1.4 Switch Stresses in the Single-Ended Isolated Converter

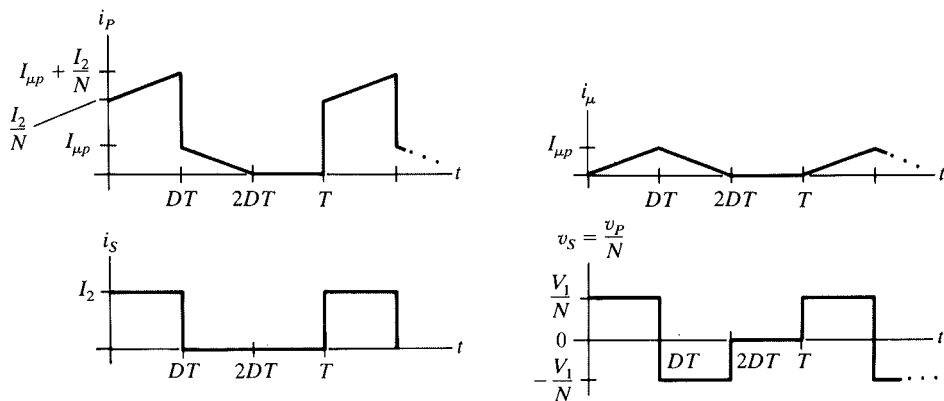
One drawback to the single-ended isolated converter is that power flows through the transformer only during the period DT . As the circuit is idle for the period $(1 - D)T$, the voltage and current stresses on the transformer, switches, and filter elements are higher than they would be if the circuit were transferring energy from input to output continuously. You can see the reason for these effects by considering the voltage and current stresses on the transistor in the circuit of Fig. 7.1.



(a)



(b)



(c)

Figure 7.4 (a) Conventional application of a clamp diode to an inductive load. (b) Extension of the circuit of (a) to provide an L_μ discharge voltage equal to $-V_1$. (c) Waveforms for the circuit of (b).

If the maximum duty ratio of the converter in Fig. 7.1 is 50%, the minimum clamp voltage, and therefore the minimum peak voltage across the transistor, is $2V_1$. If P_o is the average power flowing through the converter, the primary side current, which is the current carried by Q (ignoring i_μ), is rectangular with a 50% duty ratio and an amplitude of $2P_o/V_1$. The stress parameter of the transistor therefore is

$$V_{Qp}I_{Qp} = (2V_1)\left(\frac{2P_o}{V_1}\right) = 4P_o \quad (7.8)$$

showing that it operates with a stress factor of 4 in this circuit. Similarly, the two output diodes carry the output current and withstand twice the output voltage, so their stress factors are 2.

In general, if the maximum duty ratio is $D_{\max}$, the clamp voltage must be at least $V_1/(1 - D_{\max})$, and the current carried by the transistor is $P_o/(D_{\max} V_1)$. The stress parameter of the transistor therefore is

$$V_{Qp}I_{Qp} = P_o \frac{1}{D_{\max}(1 - D_{\max})} \quad (7.9)$$

This parameter is minimal when $D_{\max} = 0.5$. Even at the minimum value of its stress parameter, the switch must be rated at a stress factor of 4. Therefore, if we do not need isolation, we would not replace a conventional, nonisolated direct converter with one containing a transformer and a single transistor unless the output/input conversion ratio V_2/V_1 was less than 0.25 (see Fig. 6.23).

7.2 THE DOUBLE-ENDED ISOLATED FORWARD CONVERTER

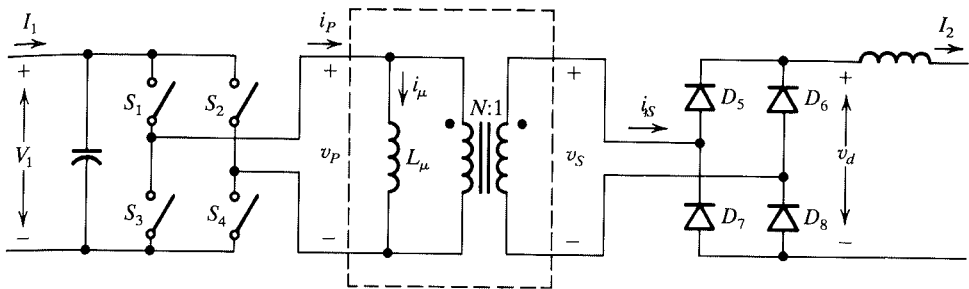
Even though the secondary voltage in the circuit of Fig. 7.4(b) is a symmetric ac waveform, we cannot use a full-wave rectifier on the output. The reason is that diodes D_1 and D_2 prevent i_p from going negative, which would be necessary if we used a full-wave rectifier circuit on the secondary voltage. The result is that the frequency of the ripple in v_d equals the switching frequency of the transistors. *Double-ended* converters permit full-wave rectification and a doubling of the fundamental ripple frequency—and also result in better utilization of transformer core material. Their disadvantage is that they are more complicated than their single-ended counterparts.

7.2.1 The Double-Ended Bridge Converter

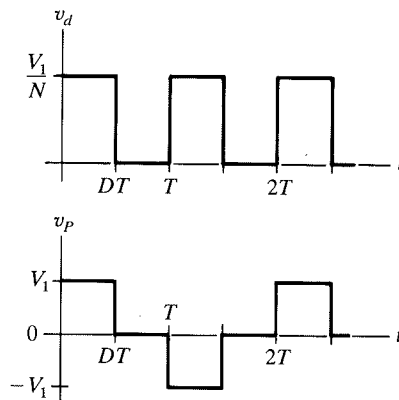
If we replace the diodes and transistors in the hybrid-bridge circuit of Fig. 7.4(b) with bilaterally conducting switches, we can use a full-wave rectifier on the output. This approach has two advantages. First, we could make the output inductor smaller, because the ripple frequency would be twice that of the single-ended circuits.

Second, the magnetizing current i_μ could change sign, thereby permitting the core material to be better used, because the flux could now vary between $\pm B_s$ instead of between 0 and $+B_s$, where B_s is the saturation flux density of the material. Such a double-ended converter is shown in Fig. 7.5(a).

The waveform of v_d for the double-ended converter is shown in Fig. 7.5(b). Here, we define the duty ratio D in terms of the period of the full-wave rectified voltage instead of the switching period of the primary side switches. If $D = 1$, each of the primary side switches would be on 50% of the time and $\langle v_d \rangle = V_1/N$. We can create the waveform of v_d by several different primary side switching sequences. They differ in the path provided for i_μ during the time between DT and T , and their applicability is constrained by the magnitude of i_μ relative to the load current reflected to the primary, I_2/N . They also result in different switch implementations. But in every case the core resets during the next interval, DT , when v_P reverses.



(a)



(b)

Figure 7.5 (a) Double-ended converter with primary side switches arranged in a bridge. (b) Rectifier output voltage v_d and primary voltage v_P .

The only function that the switches must perform during time $(1 - D)T$ is to maintain the continuity of i_μ . We do not have to apply a resetting voltage to the transformer during this time because the core resets during the next period DT when v_P reverses. We can maintain the continuity of i_μ by using any one of three switch combinations. First, we can turn off all the switches, forcing the magnetizing current to flow in the secondary winding, that is, $i_S = -Ni_\mu$. The second combination is S_1 and S_2 on, and the third combination is S_3 and S_4 on.

The first option, all the primary side switches off, is possible only if $Ni_\mu < I_2$, so that at least three secondary diodes can be on simultaneously to allow $i_S \neq I_2$. For instance, if $0 < Ni_\mu < I_2$, the diodes D_5 , D_6 , and D_7 (or, alternatively, D_6 , D_7 , and D_8) would have to be on, with I_2 circulating through D_5 and D_7 , and $i_S = -Ni_\mu$ circulating in the negative direction through D_5 and in the positive direction through D_6 . If $Ni_\mu > I_2$, then D_5 would be conducting a net negative current, which would be inconsistent with its on state. (Note that the state of D_8 during this time is ambiguous but makes no difference to the result.) The advantage of this option is that the primary switches carry only a unipolar current. But its disadvantage is that the continuity of energy stored in any transformer leakage inductance (which is not included in the figure) cannot be maintained.

The second and third options are functionally indistinguishable. In both cases the switches carry bilateral current, a path is provided to maintain continuity of primary leakage energy, and the magnetizing current i_μ will split between the primary and secondary windings. Exactly how i_μ distributes itself between the primary and secondary windings is a function of second-order parameters, such as switch and diode drops, and the relative sizes of the primary and secondary leakage inductances. In practice, we utilize both options by alternating them each cycle, as in the sequence $S_{1,4}$, $S_{4,3}$, $S_{3,2}$, and $S_{2,1}$. This ensures that each switch is subjected to the same current stress. A switch implementation suitable for this operating mode is shown in Fig. 7.6.

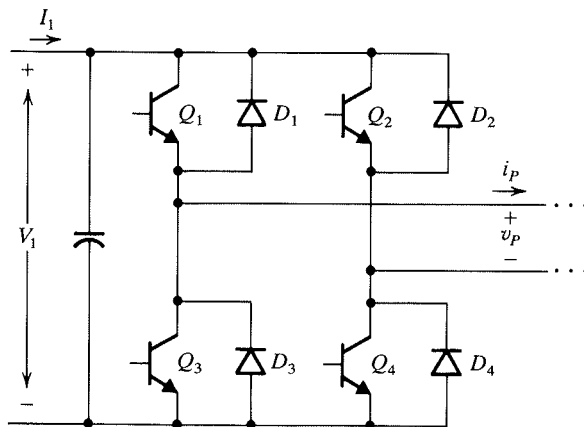


Figure 7.6 Switch implementation for the double-ended converter of Fig. 7.5(a).

EXAMPLE 7.3

Transformer Primary Current in the Double-Ended Full-Bridge Converter

Let's determine the primary current i_P in the double-ended converter of Fig. 7.5. We use the switching sequence $S_{1,4}$, $S_{4,3}$, $S_{3,2}$, and $S_{2,1}$, and we assume that i_μ circulates only on the primary side during the period $(1 - D)T$, when S_1 and S_2 or S_3 and S_4 are on. (Recall that T is the period of v_d , as shown in Fig. 7.5b.) During this time, i_μ remains constant at $I_{\mu p}$ because $v_P = 0$. A possible path for I_2 while $v_P = 0$ is the freewheeling path created if D_5 and D_7 or D_6 and D_8 are on.

At the start of the period DT at $t = 0$, we assume that $i_\mu(0) = -I_{\mu p}$. During DT , S_1 , S_4 , D_5 , and D_8 are conducting, and the primary current is

$$i_P = \frac{1}{N}I_2 - I_{\mu p} + \frac{1}{L_\mu} \int_0^t V_1 dt \quad (7.10)$$

At $t = DT$, $i_\mu = +I_{\mu p}$ (we assume operation in the periodic steady state), and S_1 turns off and S_3 turns on, maintaining a path for i_μ . With $v_P = 0$ during $(1 - D)T$, $i_\mu = I_{\mu p}$ until S_4 turns off and S_2 turns on at $t = T$. The resulting waveforms of i_μ , i_P , and i_S/N are shown in Fig. 7.7(a).

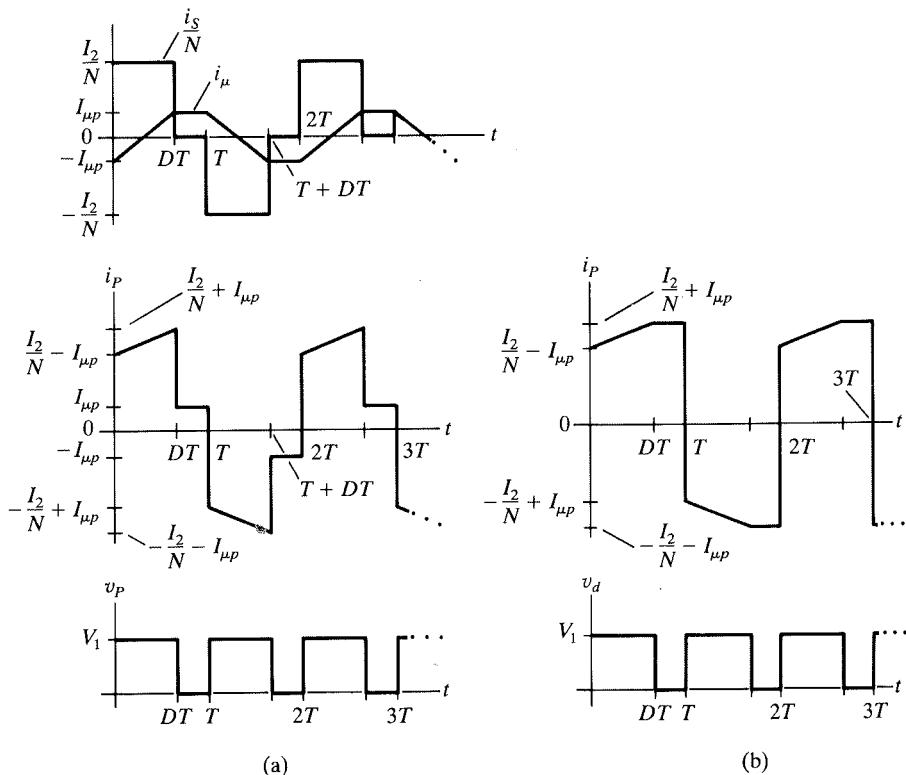


Figure 7.7 Waveforms as derived in Example 7.3 for the circuit of Fig. 7.5 or 7.6: (a) I_2 freewheeling during $(1 - D)T$; (b) D_5 and D_8 or D_6 and D_7 remaining off during $(1 - D)T$.

Although we assumed that I_2 freewheeled through D_5 and D_7 , or D_6 and D_8 , during $(1 - D)T$, the possibility exists that D_5 and D_8 remain on and that D_6 and D_7 remain off during this time, causing an additional current I_2/N to circulate through S_3 and S_4 . In this case, i_P is as shown in Fig. 7.7(b). The ambiguity present in this example disappears if the transformer model includes some leakage, for then i_S could not change instantaneously at $t = DT$, and D_5 and D_6 would remain on.

7.2.2 The Double-Ended Isolated Half-Bridge Converter

The fact that the emitters of the transistors are at three different potentials complicates the base-drive circuit for the bridge converter of Fig. 7.6. We use a center-tapped primary winding in the circuit shown in Fig. 7.8(a) to solve this problem. As a result, the emitters of both transistors connect to the same point, the negative rail of the input supply in this case. We call this converter a *push-pull* converter because of the alternating action of the two transistors. Although we show a full-wave centertapped rectifier on the secondary, a full-bridge rectifier circuit works just as well.

We control output voltage in the same way as in the full-bridge converter. But we can create the zero-voltage period $(1 - D)T$ only by turning off both primary side switches. If we ignore the magnetizing current, and the transformer has no leakage inductance, both primary winding currents are zero during this time. The transformer forces $i'_1 = i'_2$, splitting I_2 evenly between the two secondary windings, which are now shorted because both D_3 and D_4 are on. Under this condition, all the winding voltages must be zero, and therefore $v_d = 0$.

We now consider the effects of magnetizing current by modeling L_μ as an inductor across the lower primary winding of the ideal four-winding transformer. If S_1 is on, the voltage across L_μ is equal to $-V_1$, and i_μ ramps down linearly from its peak positive value $I_{\mu p}$. Because the switches are controlled to make v_P a symmetric waveform, $i_\mu = -I_{\mu p}$ at the end of the S_1 conduction period. During the S_1 conduction period, v_P is negative and D_3 is on, so $i'_1 = I_2$. As $i_2 = -i_\mu$ during this time, $i_1 = i_\mu - I_2/N$, because the ideal transformer requires that the sum of all the Ni products of the windings be zero. When S_1 turns off at DT , i_1 becomes zero, and both $i_\mu (= -I_{\mu p})$ and I_2 must be continuous. The only way for this to happen is for both D_3 and D_4 to be on. The result is that I_2 splits between the two diodes so that the difference between the two secondary currents is $i'_1 - i'_2 = Ni_\mu = -NI_{\mu p}$. Combining this expression with the constraint $i'_1 + i'_2 = I_2$ yields

$$i'_1 = (I_2 - NI_{\mu p})/2 \quad (7.11)$$

and

$$i'_2 = (I_2 + NI_{\mu p})/2 \quad (7.12)$$

Continuing this analysis results in the waveforms of Fig. 7.8(b), which were drawn for $I_{\mu p} < I_2/N$. In this case, the switch currents are unilateral, and the diodes are not required in the implementation of S_1 and S_2 . However, if $I_{\mu p} >$

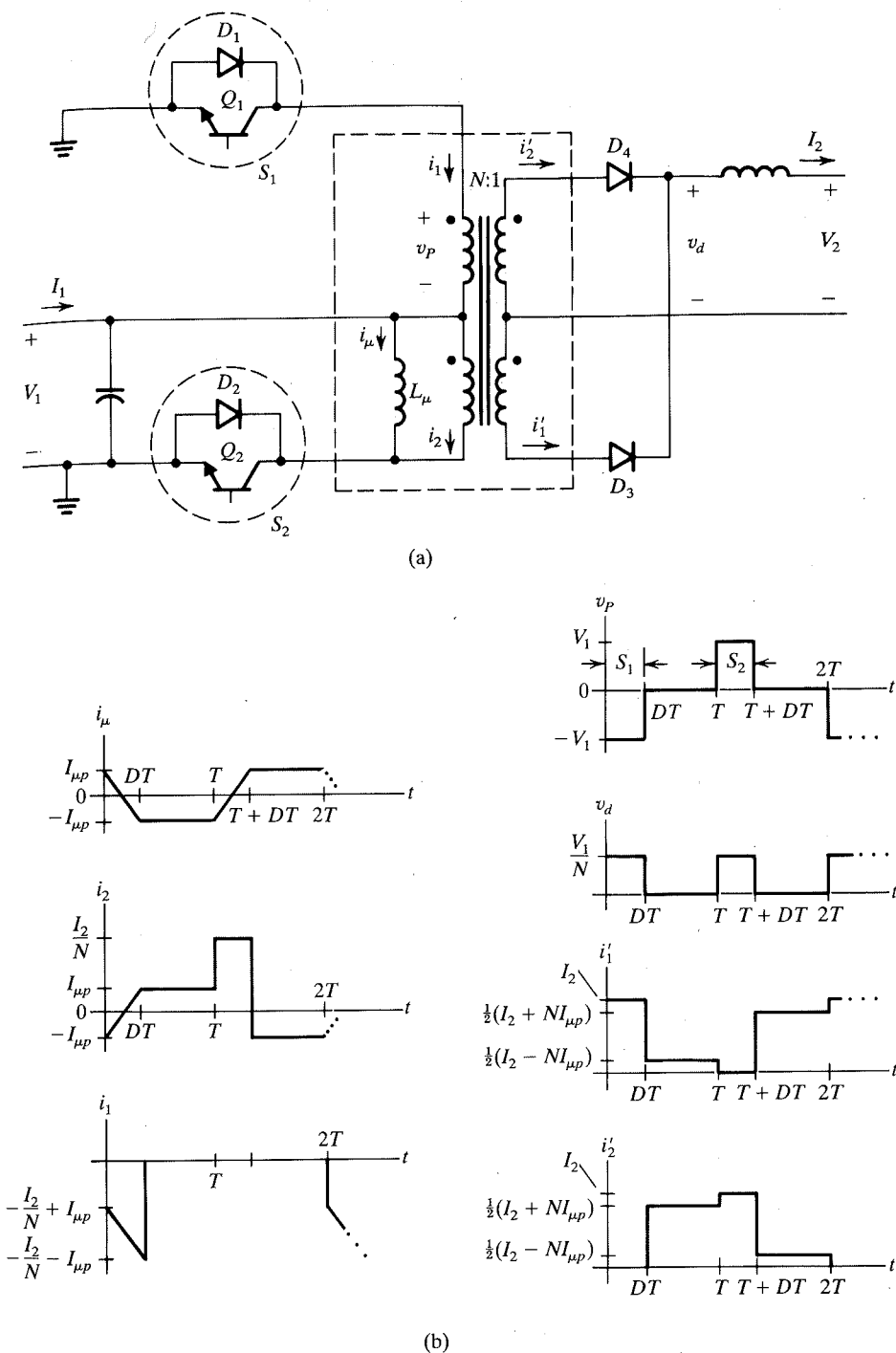


Figure 7.8 (a) A transformer coupled half-bridge, or push-pull, converter. (b) Waveforms of branch variables.

I_2/N , the switch currents are bilateral and the diodes are necessary. Note that $i_\mu(t) + i_2(t) = -i_1(t - T)$; that is, the switch currents are identical but shifted in time by T . When we recognize that L_μ is part of the transformer, we see that the circuit is symmetrical. We could model L_μ as being across any of the four windings (with a change in value to L_μ/N^2 when it is placed across a secondary winding) or even across the two primary or secondary windings. In the latter case the symmetry is more obvious.

EXAMPLE 7.4

A Current-Fed Half-Bridge Converter

So far in this chapter, we have only discussed power circuits in which the power flows from the voltage source side to the current source side of the canonical cell. In Chapter 6 we

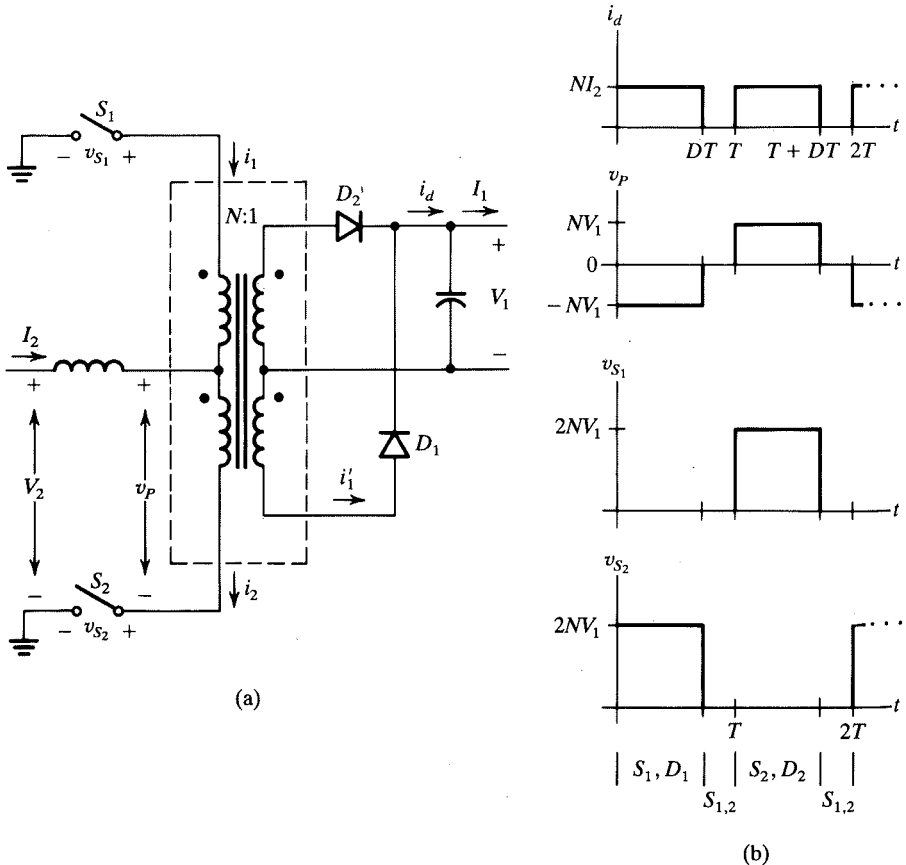


Figure 7.9 (a) Current-fed half-bridge converter of Fig. 7.8 configured to provide power flow from V_2 to V_1 ($L_\mu = \infty$). (b) Branch variables in the circuit of (a).

showed that power can flow in the other direction if we control the switches differently. To do so, in practical terms, we have to rearrange the transistors and diodes. For example, Fig. 7.9 shows the push-pull converter of Fig. 7.8(a) configured to provide for power flow from V_2 to V_1 . In this example, we assume L_μ to be infinite. This circuit is called a *current-fed half-bridge converter*. As for the up converter on which it is based, the output voltage V_1 of this converter can range from a minimum value of V_2/N (the input voltage transformed by the turns ratio) to infinity.

The operation of a current-fed converter is slightly different from the operation of a voltage-fed converter. When we turn on only one of the switches, the input current flows through the transformer to the load, and the *load voltage* determines the voltage across the transformer. To obtain an interval when power does not flow to the load (the period $(1-D)T$), we turn both S_1 and S_2 on. The input current then splits evenly between the two primary windings, just as the load current did in the secondary windings of the voltage-fed converter when both switches were off, and the transformer voltage is forced to be zero. Branch variables for the current-fed converter are shown in Fig. 7.9(b).

7.3 THE FLYBACK CONVERTER

We can easily transform the indirect converter into a circuit that provides electrical isolation because the voltage across its filter inductor has no dc component. For example, we can provide isolation in the up/down circuit of Fig. 6.18(b). We substitute a transformer having a magnetizing inductance of value $L_\mu = L$ for the filter inductor L in the circuit, as shown in Fig. 7.10(a). When Q turns on, D turns off and energy flows into L_μ . When Q turns off, D turns on and the stored magnetizing energy flows to the load through the secondary winding. Television receivers have a converter of this type with $N \ll 1$ to generate a very high voltage that drives the retrace of the beam in the picture tube. What happens visually is that the spot on the screen “flies back” to start another horizontal line. Thus the isolated indirect converter is commonly known as a *flyback converter*.

If it were not for leakage inductance, the operation of the flyback converter would be identical to that of the nonisolated indirect converter, except for the step-up or step-down created by the transformer. Primary side leakage stores energy that, in the circuit of Fig. 7.10(a), has no place to go when the transistor turns off. We usually place a snubber across the transistor to solve this problem.

Unlike the transformers used to provide isolation in the direct converter topologies discussed previously—which are transformers designed for minimum magnetizing current—the magnetizing inductance of the transformer used in a flyback converter must have a specified value in order to store energy. Such transformers are known as *energy-storage transformers*, and in the case of the flyback converter, they are called *flyback transformers*.

The Ćuk version of the indirect converter can also be isolated, as shown in Fig. 7.11. In this circuit, however, the transformer does not replace the filter inductor. Instead it operates the same as in the forward converter—it provides an impedance transformation but stores no energy.

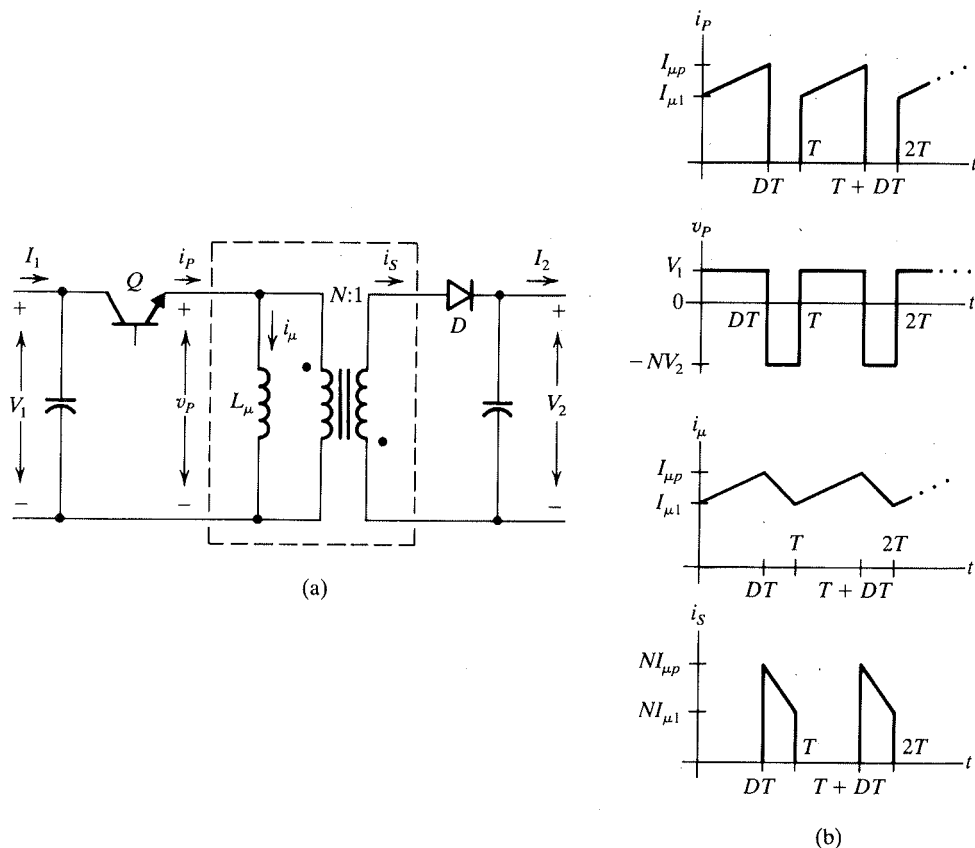


Figure 7.10 (a) The flyback converter, an isolated version of the indirect converter of Fig. 6.18(b). (b) Branch variables for the converter of (a).

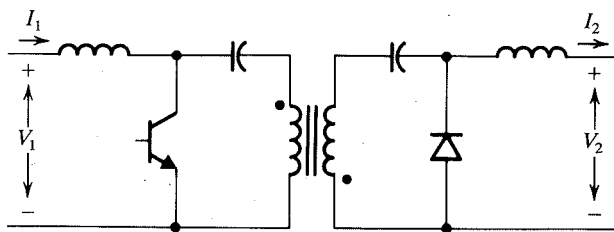


Figure 7.11 The isolated Ćuk converter.

7.4 EFFECTS OF TRANSFORMER LEAKAGE INDUCTANCE

So far in this chapter we have not considered the detailed effects of transformer leakage inductance. In this section we describe how the primary winding to secondary winding leakage inductance affects the operation of isolated converters.

7.4.1 Leakage Effects in the Single-Ended Converter

In the single-ended converter without leakage shown in Fig. 7.1, the load current immediately commutates from D_2 to D_1 when Q turns on and the voltage v_d makes a step change from zero to V_1/N . Figure 7.12(a) shows the same converter with leakage modeled as an inductor L_ℓ in series with the secondary winding. Now when the transistor turns on, the current in D_1 cannot make a step change. Instead, there is a commutation period of duration τ_{u_1} during which both diodes are on and the commuting voltage is $v_x = V_1/N$. The current in L_ℓ then increases linearly with slope V_1/NL_ℓ until it reaches the full load current I_2 , as shown in the waveforms of Fig. 7.12(b). The time τ_{u_1} is

$$\tau_{u_1} = \frac{NL_\ell I_2}{V_1} \quad (7.13)$$

Only after τ_{u_1} does D_2 turn off and allow the voltage v_d to make its step change to V_1/N .

Because $v_d = 0$ instead of V_1/N during the commutation period, the output voltage $V_2 = \langle v_d \rangle$ is less than DV_1/N , its value without leakage. Therefore, to maintain a desired output voltage over a specific range of input voltage, we must either decrease the turns ratio N or increase the maximum duty ratio. For the

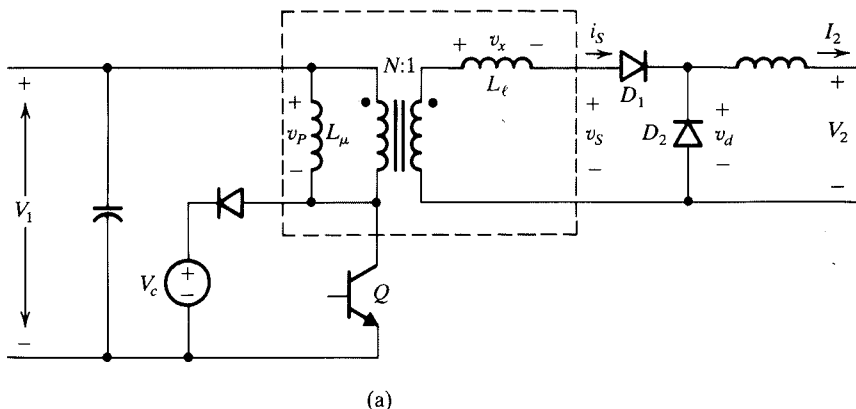


Figure 7.12 (a) A single-ended forward converter with leakage inductance.

more efficient switching regulator. By the late 1970s, the power MOSFET became economically practical and permitted efficient operation in the 200 kHz to 1 MHz range, which further reduced the size of the required switching power supply. In each case, a component improvement permitted the “breakthrough” in power electronic technology.

15.1 PRACTICAL SEMICONDUCTOR SWITCHES

Typical current and voltage waveforms for a semiconductor switching device are shown in Fig. 15.1. When the switch is on, it exhibits a forward voltage drop V_{on} , which gives a conduction loss $p = V_{on}I$ as current I flows through it. Similarly, when the switch is off, it carries a small *leakage current* that produces an off-state loss. The leakage current is usually small enough that we can neglect this off-state loss. But when the switch is a Schottky diode, or when the power circuit tries to impose a voltage across a switch that is greater than the switch’s *breakdown voltage*, the leakage current and consequent off-state loss can be significant.

A semiconductor switch also needs time to make the transition from one state to the other. During this *switching time*, $v_{sw}i_{sw} \neq 0$ and the device dissipates energy. The total energy dissipated in the turn-on and turn-off transitions, multiplied by the switching frequency, f_{sw} , gives the net *switching loss* for the device.

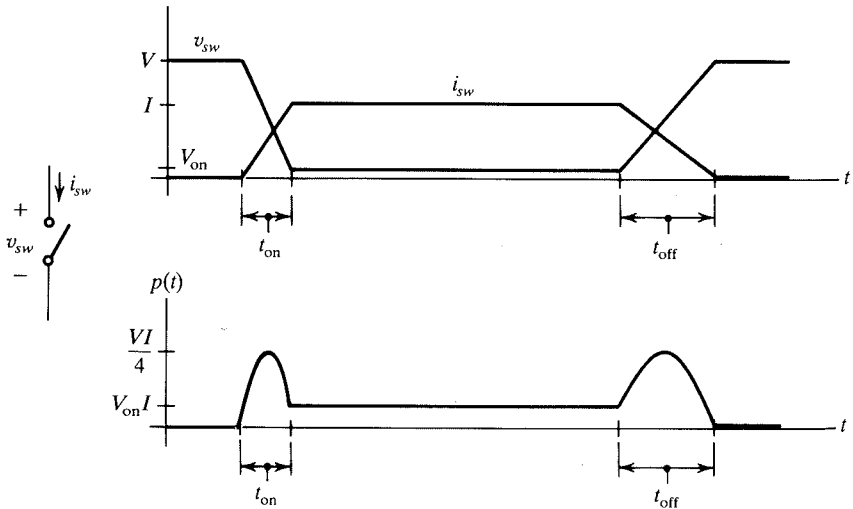


Figure 15.1 Typical voltage, current, and dissipation waveforms for a power semiconductor switch.

15.2 PRACTICAL ENERGY STORAGE ELEMENTS

We can explain some of the dissipation that occurs in real energy storage elements by the resistance of the capacitor's leads and plates and of the inductor's winding. However, calculating the loss is not simple because we must take into account the *skin effect*, which dictates where current flows in a conductor, and the *proximity effect*, which dictates how much circulating, or *eddy*, currents are induced in nearby wires or other conductive material. In general, to reduce this I^2R dissipation, we must make the energy storage element physically larger to incorporate more conductor area. Then, when the conductors are large enough for the skin and proximity effects to dominate the total loss, conductor fabrication becomes more complicated.

In addition to this I^2R dissipation, the *change* of stored energy in their magnetic or dielectric materials also causes losses in inductors and capacitors. In general, these losses are a nonlinear function of ac field strength, frequency, dc bias, and temperature. Both inductors and capacitors can be made from a variety of materials, which implies tradeoffs. One material might have less loss at a given frequency than another, but have a lower permeability (or permittivity), a lower maximum field strength, or a higher cost.

Dissipation is not the only troublesome imperfection of inductors and capacitors. Because these elements can be used as filters, we also care about whether their impedances deviate from the anticipated $j\omega L$ or $1/j\omega C$. Capacitors have a parasitic series inductance resulting from the loop formed by their plates, their leads, and their connection to other components. Inductors have a parasitic parallel capacitance caused by the proximity of two windings at different potentials. We can neglect these parasitic components for both elements if the frequency is low enough. However, at very high frequencies the "capacitor" acts like an inductor and the "inductor" acts like a capacitor. As a result, a filter having these components does not behave as expected above a certain frequency.

15.3 POWER SEMICONDUCTOR DEVICES

In this section our purpose is to outline the main power semiconductor devices available today—diodes, transistors, and thyristors—and to discuss their terminal characteristics. In Chapters 16–19 we explain in more detail the physical behavior of these devices.

At this point, however, you should know that current in a semiconductor is carried by both positive charges (holes) and negative charges (electrons). When one of these charge carriers dominates the total concentration of carriers, we call it the *majority carrier*. We call the other type of carrier the *minority carrier*. Devices whose behavior is dominated by the distribution of majority carriers (*majority*

Table 15.1 Class II Ceramic Capacitor Nomenclature

	LOW T (°C)	HIGH T (°C)	CHANGE FROM 25°C VALUE (%)
X	-55	4 +65	A ±1.0
Y	-30	5 +85	B ±1.5
Z	+10	6 +105	C ±2.2
		7 +125	D ±3.3
		8 +150	E ±4.7
			F ±7.5
			P ±10
			R ±15
			S ±22
			T +22, -33
			U +22, -56
			V +22, -82

There are two types of ceramic dielectrics, designated Class I and Class II. Class I ceramics have a dielectric constant of between 10 and 500 that is very stable (typically $\pm 3\%$) over wide ranges of voltage, time, and temperature. The dissipation factor of these materials is also low, typically 0.1% at 1 MHz. Capacitors designated NPO (or COG) are made of class I ceramic.

Class II ceramics have much higher dielectric constants than class I materials and come in two subgroups. The first has a dielectric constant that ranges from 500 to 3000 and is relatively stable over voltage, time, and temperature. The second has a relatively unstable dielectric constant ranging from 3000 to 20,000. A three-symbol code is used to specify the performance of these Class II ceramics. The first and second symbols specify the low and high limits of an operating temperature range. The third symbol specifies by how much the capacitance value at the temperature extremes will change from its value at 25°C. Table 15.1 defines this nomenclature. Two common types are X7R ($\pm 15\%$ from -55°C to $+125^\circ\text{C}$) and Z5U ($+22\%$, -56% from $+10^\circ\text{C}$ to $+85^\circ\text{C}$).

Multilayer ceramic capacitors can have very low series resistance and inductance and so are good capacitors to use at high frequencies. However, they are quite expensive compared to electrolytic capacitors, particularly for low-voltage ratings. In those cases, making the ceramic layers only as thick as required to achieve the necessary dielectric strength is mechanically difficult.

15.5 INDUCTORS AND TRANSFORMERS

Almost without exception, inductors and transformers used in power electronic circuits are custom-designed for the specific application. Seldom can these components be obtained from a manufacturer as a standard product. As a result, you

must become familiar with the materials and manufacturing techniques used in producing magnetic components, subjects covered in Chapter 20. As bases for that presentation, in Chapter 2 we discussed the behavior of transformers and their circuit models, and here we provide an overview of loss mechanisms, materials, and the common physical forms of inductors and transformers.

15.5.1 Magnetic Materials

Most magnetic components used in power circuits contain a core of high-permeability material. Use of a core results in a smaller structure than otherwise, and the core prevents magnetic fields from extending great distances from the component. If they existed, these stray fields could cause interference or induction heating of other components or structures. Two broad classes of materials are used to make cores: The most common are ferrous alloys; the others are magnetic ceramics known as *ferrites*.

The great variety of ferrous alloys are available in several forms. These alloys differ in permeabilities, loss characteristics, and mechanical workability. For low-frequency applications (below 1 kHz), iron with a low silicon content, often known as *transformer steel* is generally used. At higher frequencies other elements are added to the iron to increase its permeability and lower its loss. These more exotic magnetic steels—containing various amounts of cobalt, nickel, molybdenum and chromium—can be used to frequencies of about 20 kHz. They are very expensive relative to the silicon-iron materials. The ferrous alloys are available in either sheet or powdered form. When powdered, they can be used at frequencies higher than those cited. However, both their permeability and saturation flux density are considerably less than those of the bulk material.

Ferrites are ceramics made of ferrous oxide, zinc, and either manganese or nickel. Many proprietary types exist, and each is identified by a unique designation—for example, 4C4 or 3C6. Their maximum flux densities are about one third that of the ferrous alloys, and at temperatures above about 150°C, their magnetic characteristics deteriorate markedly. Cores made of this material are limited in size because of their brittleness. Ferrites exhibit much lower losses at high frequencies than do ferrous alloys and thus can be used at much higher frequencies. The Mn materials are useful to approximately 1 MHz, while the Ni materials are acceptable to about 10 MHz.

15.5.2 Magnetic Core Geometries

Figure 15.17 shows some of the commonly used core shapes. The ‘C,’ ‘E,’ and ‘pot’ cores are the easiest to utilize, because their windings can be made on an open bobbin. The bobbin is then slipped over the legs of the ‘C’ or ‘E’ cores or the center post of the ‘pot’ core. A toroidal core, however, must be wound by threading the wire through the core, a process that can be automated, but which is generally more expensive than winding on a bobbin.

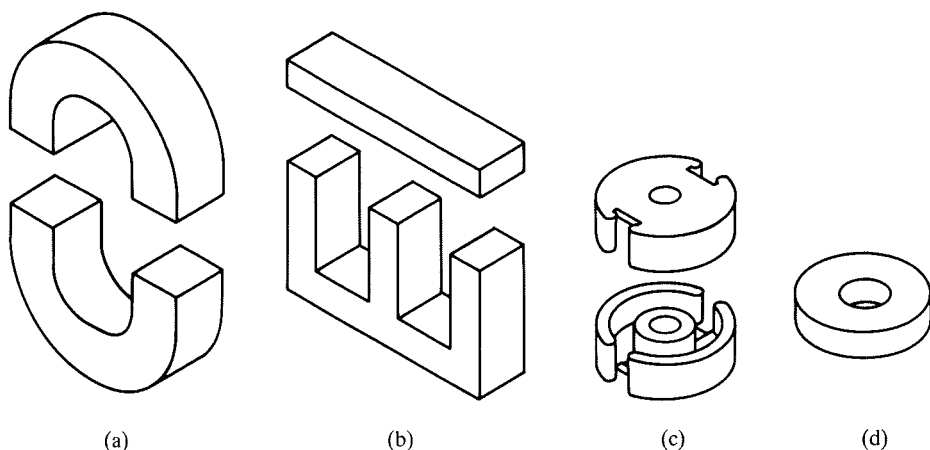


Figure 15.17 Various core geometries used in power circuits: (a) the double 'C' core; (b) the 'E' core with an 'I' closing piece; (c) the 'pot' core. (d) the toroidal core.

15.5.3 Losses in Magnetic Components

Loss in a magnetic component arises from two sources: the resistance of the copper windings; and hysteresis and eddy currents in the core. At high frequencies, the copper loss is aggravated by skin and proximity effects, but in general we can calculate it in a relatively straightforward manner. Core loss, however, is a nonlinear function of both frequency and flux level. For a particular material, core manu-

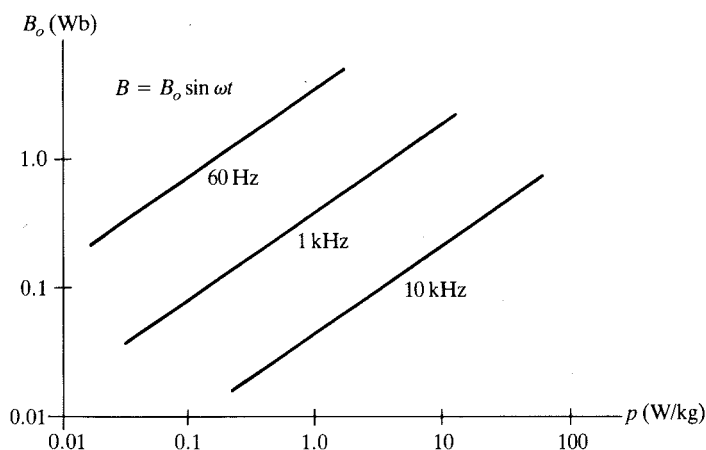


Figure 15.18 A core-loss graph for ordinary transformer steel.

5. Determine the window area (the area enclosed by the core in Fig. 20.3) required to hold this number of turns for the wire size necessary to keep the resistance below an acceptable level.
6. Check to make sure that (20.29) is satisfied for the now known length of the magnetic path.

20.2 SATURATION, HYSTERESIS, AND RESIDUAL FLUX

In Example 20.4 we showed that one benefit of a gap in a magnetic core is that it makes the value of inductance less sensitive to variations of the magnetic properties of the core material. Another advantage of a gap is that it limits the detrimental effects of *remnant*, or *residual*, magnetization. Remnant magnetization is the flux remaining in the core when the magnetic field intensity H reaches zero. Figure 20.5 shows the relationship between B and H for a typical magnetic material. As you can see, H is proportional to the current flowing in the winding, and B , through (20.16) and (20.8), is proportional to the integral of the voltage across the winding. This integral is the flux linkage λ .

Three important characteristics of the relationship are shown in Fig. 20.5. First, the path followed when the flux is increasing is not the same as that followed when the flux is decreasing. This behavior is called *hysteresis*. Second, note that neither path passes through the origin. When $H = 0$ the flux density is not zero but has a value $\pm B_r$, called the *remnant magnetization*, or *residual flux density*. When $B = 0$, the magnetic field is not zero but is equal to $\pm H_c$, a parameter called the *coercive force* of the material. Third, the slope of the path, which is the incremental

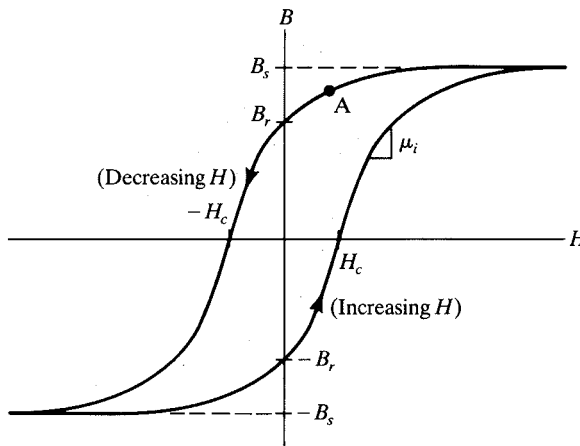


Figure 20.5 Typical B - H curve for a magnetic material.

permeability μ_i , decreases rapidly with increasing B (ultimately reaching μ_o) after some maximum $B = B_s$ is reached. This state of low incremental permeability is called *saturation* and B_s is the *saturation flux density*.

20.2.1 Saturation

Although some applications exploit the saturation characteristics of magnetic cores (for example, magnetic amplifiers, fluorescent lamp ballasts, constant voltage transformers, and self-oscillating transistor inverters), in most applications saturation is avoided. One reason for avoiding it is that in saturation the condition, (20.29), necessary to make the value of L insensitive to material properties, is often no longer satisfied because μ_c has become very small. Another reason is that in saturation, H increases very rapidly with B , resulting in large changes in current for relatively small increases in the volt-time product, λ . The final reason is that saturation reduces the coupling between windings in a transformer, because the permeance of the core is no longer much larger than that of the leakage paths.

Avoiding saturation is sometimes particularly difficult during transient circuit operation. A power circuit operating in the cyclic steady-state sequentially imposes positive and negative volt-seconds of equal value across any inductors in the circuit. Stated another way, the integral of the inductor voltage over one cycle is zero, as it must be if λ , and hence the current, is not to increase without limit. When we know the value of volt-seconds to be applied, we can design the inductor for cyclic steady-state operation so that its flux density varies from just under $+B_s$ to just over $-B_s$. However, during the first cycle when the circuit is first turned on, saturation may occur because the flux may not be starting from $-B_s$ (assuming that the first half cycle is to be positive) and therefore reaches $+B_s$ after fewer volt-seconds than anticipated. For instance, if $B(0) = 0$, only half the normal change in λ is actually available for the first cycle. This is known as the *start-up problem*.

We could solve the start-up problem by designing the inductor to support twice the normal λ . This solution, however, would require four times the maximum energy storage capability and therefore an increase of approximately four times in the physical volume of the inductor. A better solution is to control the power circuit, limiting the volt-seconds to a lower than normal value during the first cycle.

20.2.2 Residual Flux Density

The start-up problem is aggravated when the core material is characterized by a B - H curve exhibiting a high value of residual flux density. Let's consider the curve of Fig. 20.5 and assume that the last time the power circuit was operated, B and H were at point A, at which time the circuit was turned off. When the current (and therefore H) in the inductor decays to zero, B decays to B_r . The result is that if the circuit were turned on again in a positive half cycle, the volt-seconds

also very important for a transformer if the coupling between two windings is to be good. Another reason is that, for a given value of inductance and a given overall volume, the resistance of the winding is much smaller if a core is used.

20.4 TRANSFORMERS

Transformers are inductors that are coupled through a shared magnetic circuit, that is, two or more windings that link some common flux. Figure 20.12 shows a two-winding transformer. The core provides the low-reluctance magnetic circuit through which flows most of the flux generated by the windings. Transformers come in various forms and are used in power electronic circuits to perform a number of different functions. Low-frequency (50-, 60-, or 400-Hz) power transformers are usually made of laminated steel cores and function to step the line voltage up or down, to provide electric isolation, and to provide phase shifting in systems using multiphase line supplies (12-pulse rectifiers, for instance). In high-frequency converter applications, the transformer provides isolation and voltage transformation—and also stores energy when used in circuits such as the flyback converter. Manufacturers often use powdered magnetic alloys or ferrites in high-frequency transformers to control core losses. Additionally, transformers frequently provide the coupling between gate-drive or base-drive circuits and the high-power devices being driven, and are sometimes used as voltage or current sensors in feedback control systems. Transformers are frequently the most costly components in power electronic equipment.

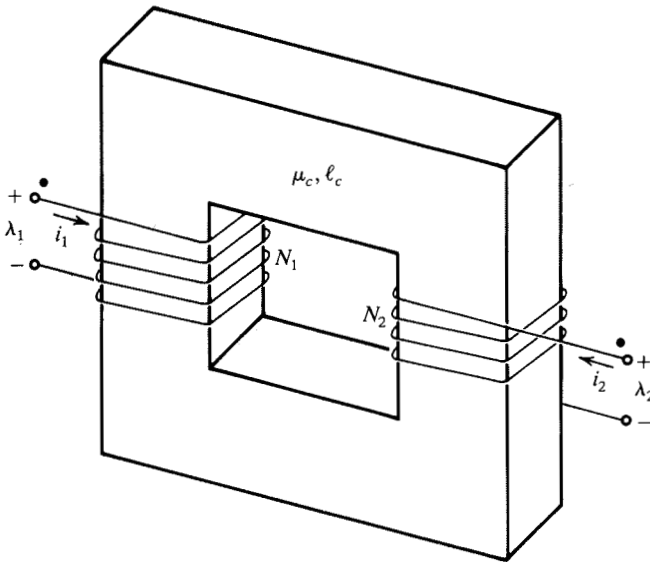


Figure 20.12 Structure of a two-winding transformer.

20.4.1 The Ideal Transformer

If the same flux is linked by both windings of a two-winding transformer—and this is the only flux they link—we say that the windings are perfectly coupled. As the flux linked by each turn is the same for both windings, the voltages induced in each turn are also identical. Moreover, the total voltage across a winding is directly proportional to its number of turns. The ratio of the two terminal voltages is therefore equal to the turns ratio, that is,

$$\frac{v_1}{v_2} = \frac{N_1}{N_2} \quad (20.48)$$

The source of the magnetic field in a transformer is the algebraic sum of the Ni products of the windings. We use a *dot convention* to indicate winding polarity. Current into the dotted ends of windings produces aiding flux. An application of the right-hand rule to Fig. 20.12 shows that current into the dotted end of either winding produces clockwise flux. We can now use Ampere's law to calculate the magnetic field intensity:

$$H = \frac{N_1 i_1 + N_2 i_2}{\ell_c} \quad (20.49)$$

If the reluctance of the magnetic path were zero, which would be the case if $\mu_c = \infty$ and there were no air gap, H would have to be zero to avoid infinite B . But H can be zero only if the Ni products for the two windings sum to zero. In this case, the ratio of the two currents is equal to the negative inverse of the ratio of the number of turns on the windings, or: 5

$$\frac{i_1}{i_2} = -\frac{N_2}{N_1} \quad (20.50)$$

The directions of the currents are opposite to each other—one flows into the dot while the other flows out.

If the terminal variables are sinusoidal, we can divide (20.48) by (20.50) to illustrate the impedance transforming property of a transformer. If $\hat{V}_1, \hat{V}_2, \dots$, are the complex amplitudes of the sinusoidal terminal variables, then:

$$\frac{\hat{V}_2}{\hat{I}_2} = Z_2 = \left(\frac{N_2}{N_1} \right)^2 \frac{\hat{V}_1}{\hat{I}_1} = N^2 Z_1 \quad (20.51)$$

where N is the turns ratio N_2/N_1 .

Equations (20.48) and (20.50) describe the behavior of an *ideal transformer*, which has the circuit symbol shown in Fig. 20.13. A real transformer differs from the ideal in three ways. First, the voltages are not exactly related by (20.48), because, owing to leakage, not all the flux linked by one winding is linked by the other. Second, finite permeability prevents the currents from being related by (20.50). A nonzero difference between them is necessary to create flux in the core. This difference is called the *magnetizing current*. Third, the relationships (20.48)

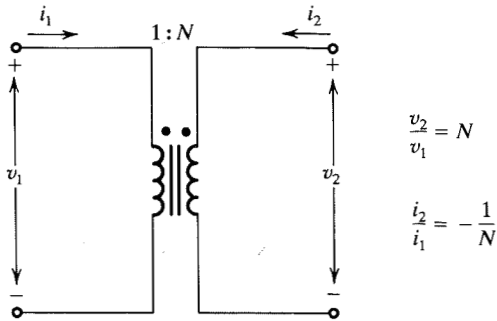


Figure 20.13 Circuit symbol that we will use and descriptive equations for the ideal transformer.

and (20.50) for the ideal transformer are not functions of frequency; that is, an ideal transformer works at dc, but a real transformer does not. In spite of these differences, the ideal transformer is extremely useful in modeling real transformers, in the same way that the ideal current source is useful in modeling transistors.

20.4.2 Magnetizing Inductance

In order for two windings to be magnetically coupled, a B must be created by one winding and linked by the other. As previously mentioned, only in the hypothetical case of infinite permeability can B exist without H and therefore without a net Ni . The closest we can come to achieving this ideal condition is to use a highly permeable core with no gap. In this case what we get from one winding, if the other is open circuited, is simply a very large—but finite—inductance called the *magnetizing inductance*. The magnetizing inductance will have one of two values, depending on which winding the inductance is being measured from, and these values are related by the turns ratio squared.

Figure 20.14(a) is a two-winding transformer with perfect coupling but a finite magnetizing inductance, and Fig. 20.14(b) is a model for this transformer. In this model we made use of the ideal transformer. The magnetizing inductance L_μ is in shunt with the ideal transformer. We can place L_μ on either side of the ideal transformer, so long as its value is correct for that side. As we said in Chapter 1, we have compromised the conventional ideal transformer symbol to obtain schematic clarity. The model in Fig. 20.14(b) for the real transformer is enclosed in a box, and the ideal transformer windings are identified by adjacent double lines.

The magnetizing inductor current I_μ is known as the *magnetizing current*. It is the current that prevents the Ni products of the real transformer's windings from summing to zero. The presence of the magnetizing inductance is the reason that a real transformer will not work with dc. (The magnetizing current would be infinite.)

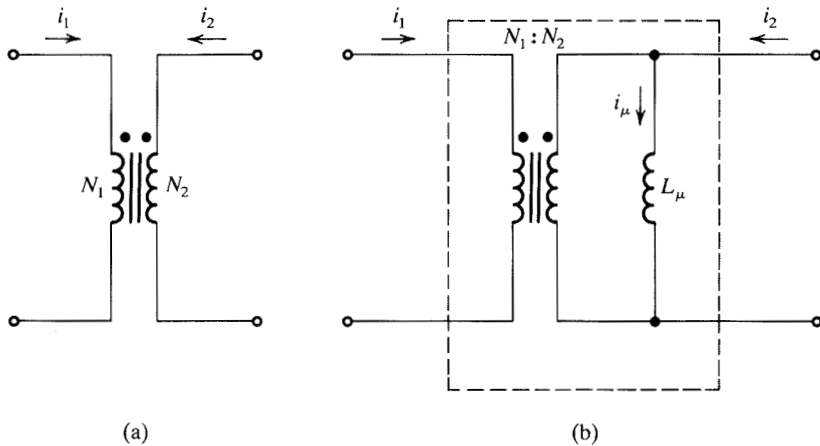


Figure 20.14 (a) A two-winding transformer with a magnetic core. (b) Model of the transformer of (a), assuming perfect coupling but finite magnetizing inductance L_μ .

EXAMPLE 20.7

Simple Use of the Transformer Model

Figure 20.15(a) shows a sinusoidal voltage source of amplitude V_1 connected to a resistor of value R through a transformer with N_1 primary and N_2 secondary turns, giving a turns ratio of $N = N_2/N_1$. As redrawn in Fig. 20.15(b) the transformer in the circuit has been

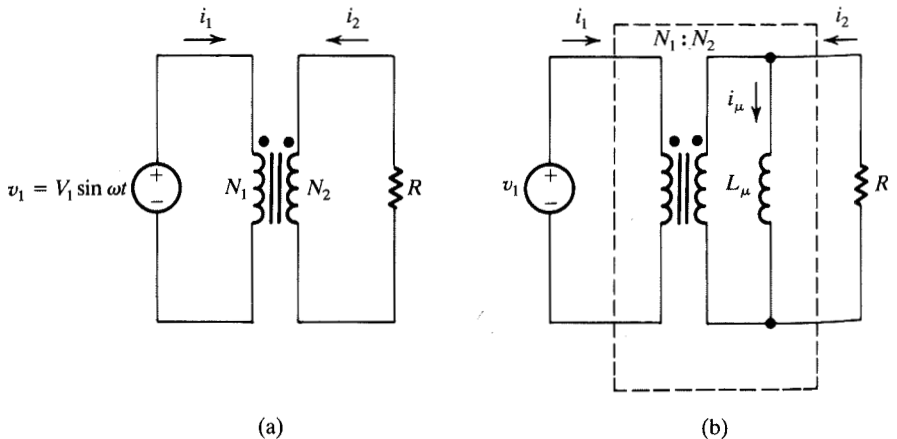


Figure 20.15 (a) Sinusoidal voltage source connected to a resistor by a transformer with N_1 primary and N_2 secondary turns. (b) Transformer in (a) replaced by a model consisting of an ideal transformer and a magnetizing inductor L_μ .

replaced by a model consisting of an ideal transformer and the magnetizing inductance L_μ on the secondary side. The core has a cross-sectional area A_c and a saturation flux density B_s . We want to know the primary current i_1 and how low ω can become before the core saturates.

The current i_1 is simply the source voltage divided by the secondary impedance transformed to the primary. That is,

$$\begin{aligned} i_1 &= \text{Im} \left[V_1 e^{j\omega t} N^2 \frac{R + j\omega L_\mu}{R(j\omega L_\mu)} \right] \\ &= \frac{N^2 V_1 \sqrt{R^2 + (\omega L_\mu)^2}}{R\omega L_\mu} \sin\left(\omega t - \tan^{-1} \frac{R}{\omega L_\mu}\right) \end{aligned} \quad (20.52)$$

We determine the lowest frequency at which this transformer can operate without saturating the core by equating (20.16) and (20.8) and assuming that $\lambda(0) = -\lambda_s$. (This assumption ignores the start-up problem.) The result is

$$\lambda_s = N_x A_c B_s \quad (20.53)$$

$$\lambda_s = -\lambda_s + \frac{1}{\omega_{\min}} \int_0^\pi V_x \sin \omega t \, d(\omega t) = \frac{V_x}{\omega_{\min}} \quad (20.54)$$

$$\omega_{\min} = \frac{V_1}{N_1 A_c B_s} \quad (20.55)$$

Note that we do not specify N_x as the primary or secondary turns in (20.53), or V_x as the primary or secondary voltage in (20.54). Only when we determine the side on which we will measure the winding voltage V_x do we specify N_x .

20.4.3 Leakage Inductance

Not all the magnetic flux created by one winding of a two-winding transformer follows the magnetic circuit and links the other winding. As discussed in Section 20.1.3, some of the flux “leaks” from the core and returns through the air. The result is that each winding links some flux that is not linked by the other, causing imperfect coupling. This effect is modeled by series *leakage inductances*, which are shown as L_{ℓ_1} and L_{ℓ_2} in Fig. 20.16. The transformer winding voltages are no longer related simply by the turns ratio, because we must now subtract the drop across L_ℓ from the terminal voltages to get the ideal transformer winding voltages.

Leakage is more important in transformers than in inductors. The reason is that leakage simply increases the value of an inductor, whereas it interferes with the basic operation of a transformer. Understanding the origin of leakage allows you to design transformers more effectively, minimizing this parameter. So, let’s consider the core and winding structure shown in Fig. 20.17. The winding is modeled as one solid conductor wrapped once around the core, whereas in reality it may be made of N turns of round wire. From the point of view of current distribution, the

Thermal Modeling and Heat Sinking

AN unfortunate consequence of our preoccupation with things electrical is that the problems of heat sinking and thermal management are frequently ignored until forced on us by sound, sight, or smell. The insatiable need to make things smaller—and the possibility of doing so by using higher frequencies and new components and materials—aggravates the problem of heat transfer, because such improvements in power densities are seldom accompanied by corresponding improvements in efficiency. Thus we are stuck with the task of getting the same heat out of a smaller volume while disallowing any increase in temperature.

The diversity of heat sources within power electronic apparatus produces a challenging cooling problem. Unlike signal processing circuits—where most heat generating components come in a common, small, and low-profile rectangular package—energy processing circuits contain components of odd shapes and orientations. Even components of the same type come in many different forms and packages. Inductors, for instance, can be small or large, round or rectangular, and with loss dominated by core or copper. Each possesses special requirements and presents a unique thermal problem. The task of integrating these parts into a reliable piece of equipment becomes as much a mechanical challenge as the circuit design was an electrical challenge.

Heat transfer occurs through three mechanisms—conduction, convection, and radiation. In conduction the heat transfer medium is stationary, and heat is transferred by the vibratory motion of atoms or molecules. Convective heat transfer occurs through mass movement—the flow of a fluid (gas or liquid) by the heat generating apparatus. The flow can be either natural or forced. In *natural convection*, the bouyancy created by temperature gradients causes the fluid to move; in a *forced convection* system, the mass flow is created by pumps or fans. Heat transfer by radiation turns the heat energy into electromagnetic radiation, which is absorbed by other elements in the environment. Except for space applications, radiation as a mechanism of heat transfer in a power electronic system is seldom of sufficient importance to justify the complexity of considering it in detail. Therefore in this chapter we focus on conduction and convection.

The material in this chapter will not give you new and novel ideas for designing heat transfer systems. The problem is too application specific to permit a discussion of that type to be of general value. Instead, we describe first the parameters governing the performance of any such system. Then we consider the modeling of both steady-state and transient thermal behavior, as applicable to power electronic systems. Some straightforward examples of specific designs will be presented to illustrate the discussions.

25.1 STATIC THERMAL MODELS

Circuit theory is the “lingua franca” of engineering for good reason. The elegance and simplicity of its canonical formulations (for example, KCL and KVL) permit complex problems to be approached in an organized way, and the insights gained through such formulations (for example, eigenvalues) are extremely valuable in predicting system behavior. Therefore many engineering problems in contexts other than electrical engineering are cast in terms of circuit models before being analyzed. One of these contexts is heat transfer.

Problems in which conduction and convection dominate radiation can be accurately transformed to electrical networks over restricted temperature ranges. Radiation is a complex mechanism that does not lend itself to linear circuit analogs. Fortunately, heat transfer from most power electronic systems is dominated by conduction and convection. If only these two mechanisms are considered, the design will be conservative, as any heat transferred through radiation will reduce the temperature of the apparatus below the design temperature. The exception is in enclosures, where radiation from hot components may be absorbed by those at lower temperatures, causing these latter components to operate at higher temperatures than anticipated. In such cases, radiation shields—shiny metal partitions—can be employed to isolate the offending or affected components. We consider here, therefore, only conduction and convection processes and the network models that represent them in the static state. In a later section we consider transient models for pulsed or short duty-cycle operation.

25.1.1 Analog Relations for the dc Case

The rate at which heat energy is transferred by conduction from a body at temperature T_1 to another at temperature T_2 is Q_{12} . It is linearly proportional to the temperature difference between the two bodies, $T_1 - T_2$, and inversely proportional to a physical parameter called the *thermal resistance* between them, R_θ :

$$Q_{12} = \frac{T_1 - T_2}{R_\theta} \quad (25.1)$$

The analogy with Ohm’s law is evident, and we can make the following assignment of variables:

$$T_{1,2} \Longleftrightarrow v_{1,2} \quad Q_{12} \Longleftrightarrow i \quad \text{and} \quad R_\theta \Longleftrightarrow R$$

Note that the analog of thermal power is i , not vi . If heat leaves body 1 only through the interface characterized by R_{θ} , then i is not only analogous to Q_{12} , but, because we are considering only steady-state conditions, it also represents the rate at which energy is being converted to heat in body 1. In the context of our interests, body 1 would be a packaged electrical network, and Q_{12} would represent the rate at which electrical energy was being converted to heat (dissipated) in the package, that is,

$$p_{\text{diss}} \longleftrightarrow i$$

The thermal management problem is to design a heat transfer system (that is, R_{θ}) that constrains ΔT (the temperature difference between component and ambient) to that value dictated by component ratings and ambient conditions.

Figure 25.1 illustrates the electrical analog for the simple two-body system just discussed. The bodies are at temperatures T_1 and T_2 and are connected thermally through the crosshatched interface, which can be characterized by a thermal resistance of value $R_{\theta 12}$. The units of T are $^{\circ}\text{C}$, and as the units of Q are watts (W), thermal resistance has the units $^{\circ}\text{C}/\text{W}$. As with electric circuits, where parallel resistances can be combined into a single equivalent resistance, parallel thermal paths, each characterizable by a thermal resistance, can be combined into an equivalent single thermal resistance.

Convection is the mechanical transport of heat by a moving fluid. The fluid (air, for instance) can move because of gravitational forces caused by density gradients, in which case the process is called natural convection. Or the fluid can be driven (perhaps by a fan), resulting in what is called forced convection. Convection is a somewhat more complex process than conduction and can be described by the relation:

$$Q_{12} = h(T, \nu)A(T_1 - T_2) \quad (25.2)$$

where ν is the fluid velocity. The parameter $h(T, \nu)$ is the *film coefficient of heat*

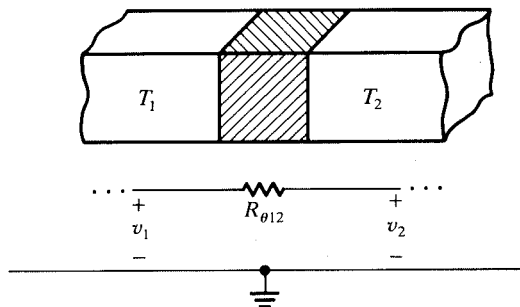


Figure 25.1 Electrical analog of simple two-body thermal system. The cross-hatched region is the thermal interface characterized by the longitudinal thermal resistance $R_{\theta 12}$.

transfer and depends on temperature and fluid velocity, and A is the cross-sectional area of the interface. Over the temperature ranges of interest (approximately -40° to $+100^\circ\text{C}$), h is fairly constant. With respect to fluid velocity, significant changes in h occur when the flow regime changes from laminar to turbulent. Within each regime, however, h is relatively constant as a function of velocity. Within these limits, the product hA may be modeled as constant, giving to (25.2) the same form as (25.1), with $R_{\theta 12} = 1/hA$. Thus the electrical analog shown in Fig. 25.1 is appropriate for representing convective as well as conductive heat transfer.

25.1.2 Thermal Resistance

A thermal resistance can be used to model both conductive and convective heat transfer, as we have just shown. The physics governing thermal conduction is much like that for electrical conduction, and the thermal resistance or conductance can be described in terms of parameters abstracted from the physics of the process (for example, conductivity, and geometry). In fact, thermal and electrical conductivity are intimately related by the *Wiedeman–Franz law*—materials of high electrical conductivity are also good thermal conductors. Convection, however, depends on parameters that are not so easily abstracted. For instance, while conductivity can be adequately described in terms of material type and temperature, h is a function not only of these parameters but also of velocity, surface characteristics, and geometry. The latter parameter often determines the *Reynolds number*, a dimensionless number that indicates whether laminar or turbulent flow will occur in a particular situation. Furthermore, the geometry of the convective part of the system is invariably complex, consisting quite often of a finned aluminum extrusion. Thus the equivalent thermal resistance model for convective transfer from a specific heat sink type in a variety of environments (for example, forced or natural convection) is tabulated by the manufacturer. For this reason, the following discussion is directed at determining the equivalent thermal resistance for those parts of the system where heat transfer is by conduction.

The analog of electrical resistivity ($\Omega\text{-m}$, or more commonly, $\Omega\text{-cm}$) is thermal resistivity, ρ_θ , in units of $^\circ\text{C-cm/W}$ (or $^\circ\text{C-in./W}$ in a field noted for its lack of concern about mixed units). In terms of ρ_θ and physical dimensions, the longitudinal thermal resistance of a piece of material of cross-sectional area A and length l is

$$R_\theta = \frac{\rho_\theta l}{A} \quad (25.3)$$

The thermal resistivities of various materials used in heat transfer paths in electronic equipment are shown in Table 25.1. Mylar, and less commonly mica, is used to provide electrical isolation between electrically hot components (for example, the semiconductor device package and the heat sink). Mica has a much higher dielectric strength, is more impervious to mechanical puncture, and can be cleaved to produce thinner sheets than mylar—but is more expensive. Beryllia and alumina,

Table 25.1 Thermal Resistivities of Materials Used in Electronic Equipment

MATERIAL	RESISTIVITY ($^{\circ}\text{C}\cdot\text{cm}/\text{W}$)
Still air	3050
Mylar	635
Silicone grease	520
Mica	150
Filled silicone grease	130
Alumina (Al_2O_3)	6.0
Silicon	1.2
Beryllia (BeO)	1.0
Aluminum Nitride (AlN)	0.64
Aluminum	0.48
Copper	0.25

and recently aluminum nitride (AlN), are also used to provide electrical isolation, most frequently within device packages.

Silicone grease impregnated with metallic oxides, such as ZnO_2 , is used to fill imperfections such as scratches on mating surfaces in a heat transfer path—between the bottom of a device package and the top of a heat sink, for instance. The need to fill these voids with something besides air is apparent from the table. Filled silicone grease is also referred to as *thermal grease* or *thermal compound* (or “goop” for reasons that become clear when you use it). Anodization is frequently used to create an attractive or black surface on aluminum components. Since the resulting oxide is very thin, its contribution to the thermal resistance of a path is negligible. Although the oxide is a good insulator, it is unwise to rely on it in lieu of mica or mylar for providing electrical isolation between surfaces. Note that the ratio of ρ_θ for copper and aluminum is the same as that for their electrical resistivities. This result relates to the similarity of the physics of both heat and charge transport in metals and is described by the Wiedemann–Franz law.

EXAMPLE 25.1

A Calculation Using an Electrical Analog

Figure 25.2(a) shows a resistor embedded in a block of aluminum. What is the temperature of the resistor if it is dissipating 50 W and the ambient temperature is $T_A = 75^{\circ}\text{C}$?

As the length of the block (10 cm) is much longer than the radius of the resistor (3 mm), we can assume that the power is dissipated uniformly in a plane centered between the two heat sinks, which are at temperature T_A . This assumption says that the detailed pattern of heat flow in the vicinity of the resistor is unimportant. The resulting analog circuit model

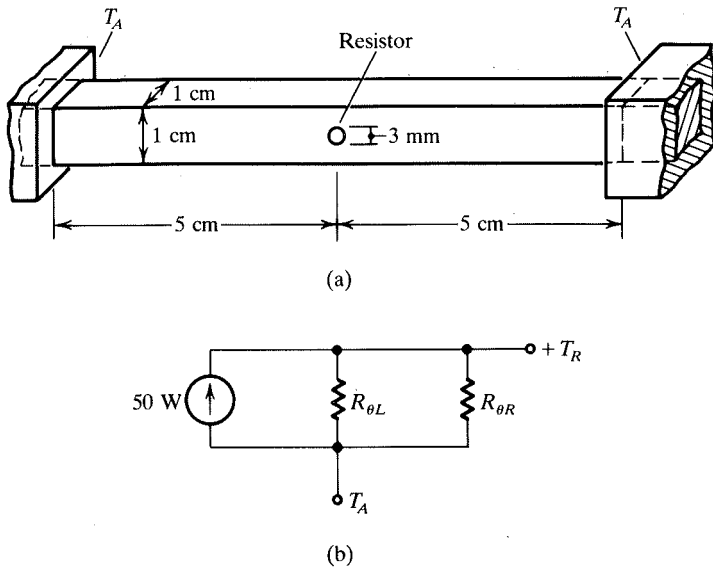


Figure 25.2 (a) A thermal system consisting of a resistor embedded in the center of an aluminum block. (b) The electric circuit analog for the thermal system of (a).

is shown in Fig. 25.2(b), where $R_{\theta L}$ and $R_{\theta R}$ are the thermal resistances of the aluminum bar to the left and right of center. The value of these resistances are

$$R_{\theta L} = R_{\theta R} = \frac{(0.48)(5)}{1} = 2.4^{\circ}\text{C/W} \tag{25.4}$$

At a dissipated power of 50 W, the temperature of the resistor, T_R , is

$$T_R = 75 + 50\left(\frac{2.4}{2}\right) = 135^{\circ}\text{C}$$

25.2 THERMAL INTERFACES

A critical part of any heat transfer system is the interface between mechanical components in the thermal path. Some issues related to these interfaces in the context of our application were raised in the previous section. The geometry of most interfaces can be modeled as two parallel planes with material of a specific thermal resistivity between. If the material is of thickness δ and of area A , the thermal resistance of the interface between the planes is

$$R_{\theta i} = \frac{\rho_{\theta} \delta}{A} \tag{25.5}$$

Consider now that the interface is between a TO-3 package and a flat plate, and consists of 0.05 mm mylar. The mating surface area of a TO-3 package is ap-

proximately 5.0 cm^2 , giving a thermal resistance between the case and the sink of:

$$R_{\theta CS} = \frac{(635)(0.005)}{5.0} = 0.64^\circ\text{C/W} \quad (25.6)$$

Thus the difference between the case and sink temperatures increases by 0.64°C for each watt of thermal power being transported across the interface. A dissipation of 25 W is not unusual for a device in a TO-3 package, giving a temperature rise of 16°C across just the mylar interface. This amount is a lot and illustrates the price paid for requiring isolation.

EXAMPLE 25.2

A Thermal System

The physical structure depicted in Fig. 25.3 is typical of the thermal system that results from mounting semiconductor devices. The silicon device itself is bonded to the header using solder or epoxy, the header is made part of a package that is mounted to a heat sink (perhaps with some intervening insulating material), and the heat sink is thermally connected to the ambient environment, generally through free or forced (fan) convection. An overly detailed model for this system is shown in Fig. 25.4(a), where each part of the system is explicitly represented by its equivalent thermal resistance. The model also shows the relationship between certain node voltages and the temperatures they represent. The current source represents the rate at which electrical energy is dissipated in the device, P_{diss} . The physical location within the device of this dissipation is the node to which the source is connected, the junction in this case.

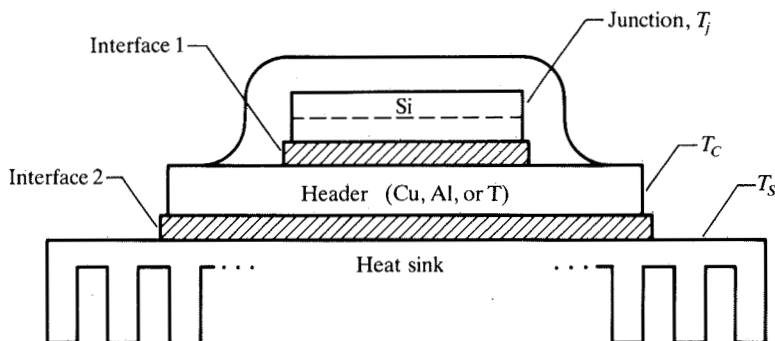


Figure 25.3 Typical mechanical structure used for mounting a semiconductor device.

Some of the thermal resistances shown in Fig. 25.4(a) are so small relative to others that they can be neglected. Because the header is made of copper or aluminum, its vertical thermal resistance is negligible ($\approx 0.03^\circ\text{C/W}$ for an aluminum TO-3 package), as is that of the thermal grease (assuming that it is applied properly, which it often is not!). Others of the identified resistances are frequently lumped together, such as those for the silicon and bonding material, which are generally inaccessible to the circuit designer. Implicit in the element $R_{\theta SA}$ are the thermal resistance of the sink extrusion between the region on which

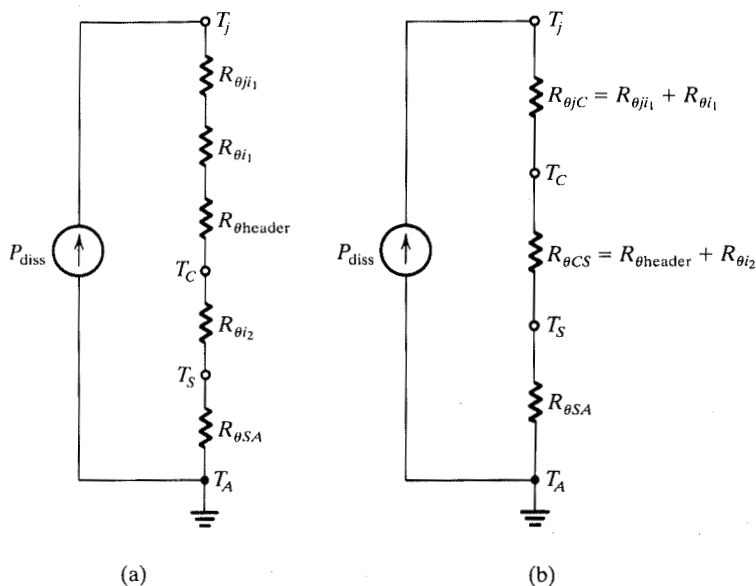


Figure 25.4 (a) A static thermal model for the structure of Fig. 25.3. (b) Simplified model of the circuit in (a).

the package is mounted and the surfaces from which heat is being convected away, and the thermal resistance representing the convection process. Figure 25.4(b) shows the simplified model.

The variable of interest in the models of Fig. 25.4 is T_j , the “junction” temperature of the device. The term “junction” is used rather loosely to represent in lumped form the source of heat in the device. In reality, this source seldom exists as a simple plane. In the MOSFET it is not a junction at all. Nevertheless the term persists, and manufacturers determine $R_{\theta jC}$ empirically, which takes into account the actual geometry of the heat-producing region. To determine T_j , we need to know P_{diss} as well as the thermal resistances between the junction and the ambient environment.

The dissipation in the device is a function of its electrical environment (for example, its current, voltage, and switching loci). For purposes of this example we assume that these calculations have been made for our device and that the result is $P_{diss} = 25$ W. The physical configuration is a TO-3 package mounted on an extruded, finned, free convection-cooled sink without any insulating interface but with thermal grease. Typical values of the thermal resistances are $R_{\theta jC} = 0.75$, $R_{\theta CS} = 0.12$, and $R_{\theta SA} = 1.8$, all units being $^{\circ}\text{C}/\text{W}$. The last parameter needed is the ambient temperature, which is not “room temperature,” but the temperature of the air in the vicinity of the sink. We take it to be 40°C . We determine the temperature drop between nodes in the model by using Ohm’s law, obtaining

$$T_S = 40 + (25)(1.8) = 85^{\circ}\text{C}$$

$$T_C = 85 + (25)(0.12) = 88^{\circ}\text{C}$$

$$T_j = 88 + (25)(0.75) = 107^{\circ}\text{C}$$

Is this an adequate design? The answer depends on the type of device being cooled. If it were a bipolar or MOSFET transistor, the design gives a good margin between predicted junction temperature and typical maximum limits of 150°C. For a thyristor the design is marginal. What would be the influence of an isolating interface between package and sink?

25.2.1 Practical Interfaces

Mechanical interfaces are not parallel in practice. They contain surface imperfections, such as scratches, and a characteristic called *run-out*. Run-out is the maximum deviation from flatness that a surface exhibits over a specified lateral distance. It is measured in (linear dimension)/(linear dimension), for example, cm/cm. A standard aluminum extrusion exhibits run-out that is typically 0.008 cm/cm. Both of these surface characteristics degrade the thermal performance of an interface.

The use of thermal grease has already been mentioned. It is designed to reduce the degrading effects of surface scratches and other small imperfections but is not designed to remedy the effects of run-out. Although our primary focus is on basic principles, thermal grease is so frequently misused that a brief departure from “principle” to “practice” is justified. The problem arises from a belief that if a little is good, a lot is better. However, the art of applying thermal grease is much like that of watering plants—too much and it’s dead. Silicone grease is highly viscous and refuses to “squish out” when squeezed between header and sink by mounting hardware. In such cases, a thin layer of grease can remain in the interface, giving rise to a significant thermal resistance that was not anticipated in the thermal design. The grease should be applied sparingly, and then wiped off, removing almost all traces. Thermal grease oozing from under device packages is a sign of poor construction and potential thermal problems.

Run-out is usually not under the design engineer’s control. Therefore we must determine the run-out characteristics of our components, and make proper allowance for them. As an example, Fig. 25.5 shows the resulting thermal resistance as a function of the sum of both header and sink run-out for several common device packages.

25.2.2 The Convective Interface

Even though we showed that both conduction and convection processes could be modeled by similar electrical analogs, our discussion so far has focussed on conduction interfaces. The fact of the matter, however, is that all conduction leads to a convective interface. Heat is removed from a conventional finned sink by air flowing over the fins. A more sophisticated system incorporating a heat exchanger probably uses a liquid to move the heat from one place to another. These are convective processes. As mentioned earlier, the physics governing these processes is beyond our scope here. But a short discussion of the application of finned sinks is helpful.

The critical issue in the application of finned sinks is to ensure that air flow through the fins is turbulent rather than laminar. Laminar flow, as the name implies,

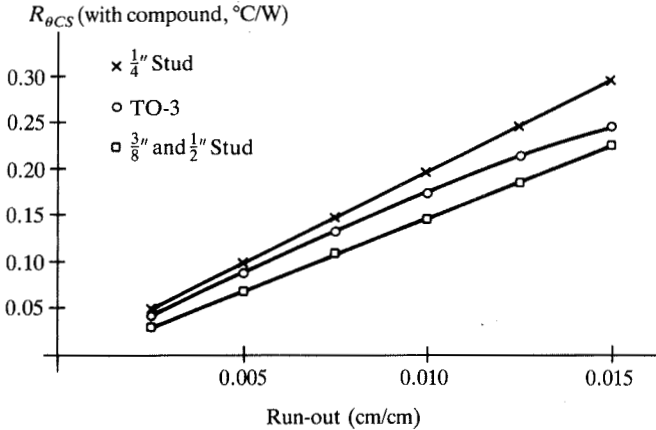


Figure 25.5 Thermal resistance of the interface between various device packages and sink surface as a function of run-out. (Data from W. E. Goldman, "An Introduction to the Art of Heat Sinking," *Electronic Packaging and Production*, July 1966.)

is the flow of a fluid in such a way that strata can be defined; that is, all flow is in one direction. The important feature of laminar flow is that no mixing of strata occurs. Turbulent flow, on the other hand, causes considerable mixing. Without such mixing, the particular stratum of fluid in contact with the fin would remain in contact with it for its entire length, resulting in a very low value for the film coefficient of heat transfer h , discussed in Section 25.1.1. Stated another way, the boundary layer next to the fin surface remains intact and virtually stationary in laminar flow, preventing efficient heat transfer from the fin to the moving stream of air (or other fluid).

The boundary between laminar and turbulent fluid flow is a function of many variables. But fin geometry and flow rate are the critical ones for our application. The essential influences of these two parameters quickly become evident to anyone who has canoed in rapids. A stream moving "slowly enough" will be smooth, a reflection of laminar flow. Increase the flow and eddies, holes, standing waves, and other hideous 3-dimensional phenomena arise to humble us. This same turbulently flowing stream, when constricted to a narrow passage (known as a "chute") becomes instantly laminar and glass smooth, even though the canoe shoots through it like a rocket. The salient feature of the chute is that it is not wide enough to permit turbulence. So it is with fins too closely spaced and turbulence won't occur.

The relationship among fin spacing, flow rate, and the onset of turbulence is given by the *Reynolds Number*. A high Reynolds Number is characteristic of turbulent flow; a low number is characteristic of laminar flow. The Reynolds Number

Re for a fluid flowing at velocity ν through a channel of width w is

$$Re = \frac{\rho \nu w}{\eta}$$

where ρ is the fluid density, and η is its coefficient of viscosity. This expression shows that fluid flowing in a wider channel will enter the turbulent flow regime at a lower velocity than that flowing through a narrower channel. The point here is that, like thermal compound, more is not necessarily better. Because of reduced turbulence and flow rate, a sink with many closely spaced fins and a large surface area could produce poorer thermal performance than one with fewer, but more widely spaced fins and a smaller surface area.

Although we have been using the context of semiconductor heat sinks for this discussion, it is equally appropriate to the cooling geometry associated with other components. Closely spaced parts impede proper convective flow for the same reasons that too closely spaced fins do.

25.3 TRANSIENT THERMAL MODELS

So far our discussion and models have been limited to systems in which both the energy being dissipated and the temperatures within the system are constant. Our models do not represent the thermal processes associated with start-up, where dissipation may be constant, but temperatures are climbing. Or pulsed operation, where temperatures may be constant, but dissipation is not. The latter situation is the more important, for under such conditions the permissible instantaneous dissipation can be much higher than predicted by static thermal models. Essentially the heat capacity of components or their constituent parts creates a low-pass filter, which in the limit of small bandwidth only responds to the dc in $p_{\text{diss}}(t)$.

Heat capacity is a measure of the energy required to raise the temperature of a mass by a specific amount. In SI units it is specifically the energy in joules required to raise one kilogram of the material one centigrade degree. It has the units $\text{J}/^{\circ}\text{C}\cdot\text{kg}$. Water has one of the largest thermal capacities of any fluid at room temperature: $4.2 \times 10^3 \text{ J}/^{\circ}\text{C}\cdot\text{kg}$.

Masses in a thermal system, then, constitute thermal energy storage devices, and thermal systems containing mass will exhibit dynamic behavior. Since thermal power is Q , the analog of thermal capacitance is electrical capacitance in the electric circuit model for heat transfer. In its simplest form, then, the dynamic model for a mass being supplied with heat energy is an RC circuit, as shown in Fig. 25.6. If the heat source is constant, the final temperature is analogous to the final voltage across the capacitor, that is, $R_{\theta}Q$. Previously we have dealt with the steady-state solutions to such thermal systems. Now we are concerned with the transients leading to these steady states.

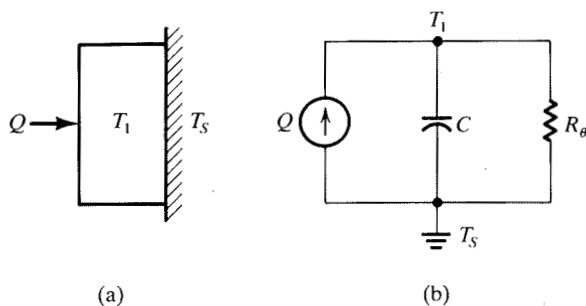


Figure 25.6 (a) A simple thermal system consisting of a mass at temperature T_1 being supplied heat Q and in contact with a sink at temperature T_S . (b) A single lump dynamic model for the system shown in (a).

25.3.1 A Lumped Model and Transient Thermal Impedance

Energy in a mass is stored in a continuum. But as is done for a transmission line, this continuum system is modeled by a series of lumped electrical elements. Consider, for instance, the mass of Fig. 25.6 with a source of heat energy applied at one end. The mass can be broken into an arbitrary number of sections, each assumed to be at a uniform temperature. These sections are each characterized by a heat capacity and are interconnected through a thermal resistance. This thermal resistance is that of the mass section between the two interfaces with other sections. This multilump model of the mass of Fig. 25.6 is shown in Fig. 25.7. The number of lumps, 5, was chosen arbitrarily.

When a continuous system is modeled by lumps, each lump displays the aggregate behavior of the physical piece of the system it represents. The model of Fig. 25.7 has been constructed so that the node voltages represent the section tem-

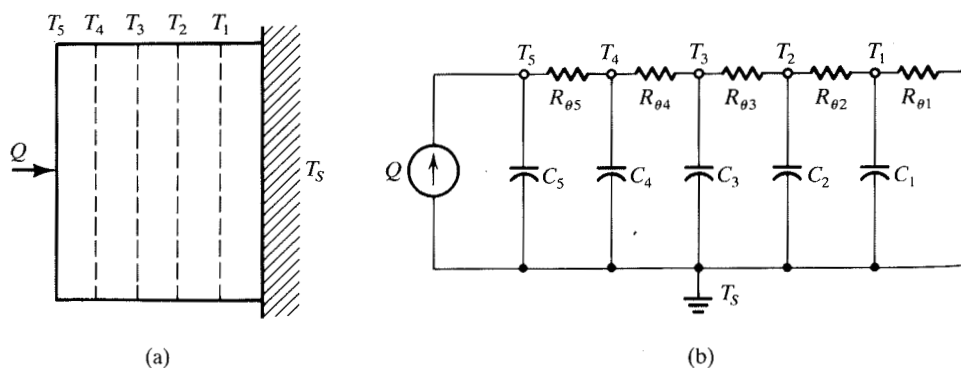


Figure 25.7 (a) The thermal system of Fig. 25.6(a) divided into five "lumps." (b) The lumped electrical analog model for the thermal system of (a).