Efficiency can also be important for off-line converters. Since typical house circuits are limited to 20A, if a converter is very inefficient, it may not be possible to run its load from a normal outlet: so much power is wasted in the converter that not enough can be delivered to the load without tripping the breaker.

Modules

It is a sorry fact that one portion of the power supply industry routinely publishes figures that are misleading, at best. Modules (as opposed to VRMs, voltage regulation modules, which are used exclusively on computer motherboards) are small converters sold in a low profile housing, typically meant to be soldered into a PCB. Now while it is frequent practice in the power supply industry to cite efficiency at whatever output load maximizes efficiency (though vendors ought to state something like "efficiencies as high as... "), the module industry cites efficiencies for the module alone. That is, their literature fails to acknowledge that many applications, if not most, will require additional components to make a functional power conversion system—in particular EMI filters. Unless you are in the know, you will be sorely surprised when you buy a module, discover that additional pieces are needed, and then discover that these additional pieces cause a big drop in the wonderful efficiency you were promised. All module vendors cite efficiency in these misleading terms because any manufacturer who broke from the pack would be at a disadvantage relative to the others in the efficiency spec on the data sheet. Maybe an industry-wide standard could be set up to fix this problem?

90% Is Doing Great!

Although no general rules can be given, here are some ideas that will help you in deciding how hard it will be to meet a particular efficiency requirement.

1. As the output voltage of a converter goes below 5V, the losses in an output diode increase as a proportion of total power (because the forward drop of the diode is always about the same voltage); if you must have > 80% efficiency at < 5V out, you're probably going to need synchronous rectification.
2. As discussed in more detail below, at low power (< 1–2W), losses in the IC supply current and gate drive current may dominate the efficiency. At this power level, 70% efficiency is doing a good job. For maximum efficiency, you will want to use a CMOS PWM, and a diode instead of a synchronous rectifier.
3. Higher efficiency almost always requires larger magnetics.
4. Very high efficiencies can be obtained in converters that have high voltage at both input and output, since the currents will be lower for a given power level; converter losses are proportional to either I or I^2.
5. Almost no converter in the low to medium power range is going to exceed about 95% efficiency. As a conceptual aid, let's suppose that you build a 100W input power converter. If this converter is 80% efficient, then its output power is 80W, and its internal losses are 20W. Increasing this efficiency by 2%, to 82%, involves getting 82W out—in other words, saving 2W out of 20W, which is 10%. On the other hand, suppose that the converter is already 90% efficient, so that its output power is 90W, and its internal losses are 10W. Now increasing the efficiency by

2%, to 92%, involves getting 92W out—or in other words saving 2W out of 10W, 20%. It is clear that saving 10% of the losses is much easier than saving 20% of the losses: increasing the efficiency by 2% becomes radically harder as the efficiency starts to climb over 90%. The moral of the story is that achieving a 90% efficient converter is doing great; if higher efficiency must be achieved, heroic efforts may be required.

Example Calculation 1

It is possible to get a pretty good estimate of the efficiency of a converter before you build it. Indeed, if a high efficiency is required, you certainly need to make such an estimate as part of the topology selection process; choosing the wrong topology may result in large costs (and headaches) later, when it becomes necessary to try to boost the efficiency. As an example, let's analyze the efficiency of a 10W output, discontinuous conduction mode, isolated flyback converter (Figure 8.1). In fact, we can use the one for which we designed the transformer in Chapter 5 on magnetics, since we already calculated the losses in the transformer there as 150mW.

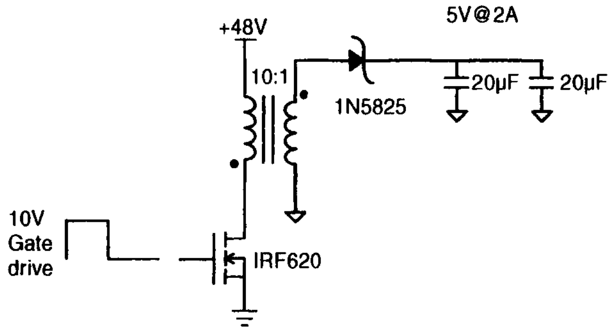


Figure 8.1 Power stage of a discontinuous isolated flyback converter.

The parameters assumed in Chapter 5 were as follows: input, 48VDC with no variation; duty cycle, 45% at this voltage; switching frequency, 250kHz. (When the forward drop of the diode is included, the duty cycle will turn out to be slightly higher than 45%; this difference has no significant effect on the transformer losses.) Recall that the design of the transformer didn't require us to specify the output voltage or the turns ratio; here we have set the output to 5V at 2A, which together with a turns ratio of 10 : 1 will ensure discontinuous conduction, as calculated below.

Let's start by recalculating the duty cycle, DC. At 2A, the forward drop of the schottky is about 300mV. This means that the power actually being delivered by the input is (5V + 0.3V)2A = 10.6W. Using the formula in Chapter 5, we write

$$DC = \frac{\sqrt{2LfP}}{V} = \frac{\sqrt{2 \times 93\mu\text{H} \times 250\text{kHz} \times 10.6\text{W}}}{48\text{V}} = 0.463$$

At this duty cycle, the peak current in the primary is

$$I_{\text{pk}} = \frac{V}{L} \times \text{DC} \times T = \frac{48\text{V}}{93\mu\text{H}} \times 0.463 \times 4\mu\text{s} = 0.956\text{A}$$

and the RMS current is

$$I_{RMS} = I_{pk} \times \sqrt{\frac{DC}{3}} = 0.956 \times \sqrt{\frac{0.463}{3}} = 0.376A$$

Since we already know the losses in the transformer, the first item to calculate is the losses in the MOSFET. These arise from three main sources, as discussed in Chapter 3: conduction losses, which are $P_c = I_{RMS}^2 R_{DS,on}$; switching losses, which are $P_{SW} = I_{pk} V_{pk} t_s f_s / 2$; and gate charge losses, which are $P_g = Q_g V f_s$.

As discussed in Chapter 3, the $R_{DS,on}$ of a MOSFET is temperature dependent. Let's assume that through a self-consistent calculation of the thermal characteristics of the supply, we have found that the temperature of the MOSFET die is 60°C. The IRF620 MOSFET [1] has a guaranteed maximum $R_{DS,on}$ of 800mΩ at 10V gate drive voltage and 25°C. For this and other data, see Figure 8.2.

Practical Note A good approximation to the temperature dependence of the $R_{DS,on}$ of a MOSFET is:

$$R(T) = R(25°C) \times (1.007^{T-25°C})$$

Using this Practical Note, we find that at 60°C and with 10V of gate drive voltage, the MOSFET has an $R_{DS,on}$ of $800m\Omega \times 1.007^{35} = 1.0\Omega$. The MOSFET's conduction losses are thus $P_c = (0.376A)^2 \times 1\Omega = 141mW$. Assuming that the drain switches in 50ns, switching losses are $P_{SW} = (0.956A \times 48V \times 50ns \times 250kHz)/2 = 287mW$. Typical gate charge at 10V and 48V drain–source voltage for the IRF620 from Figure 8.2 is about 9nC, so that gate charge losses are $P_g = 9nC \times 10V \times 250kHz = 22mW$. Total losses in the MOSFET are thus $P_{tot} = 141mW + 287mW + 22mW = 450mW$.

The diode losses [2] are set by the diode forward voltage and the diode current. We need to be just a little careful here—although the current in the calculation is the diode average current (2A, here), the forward voltage used should be the V_f at the average current during the conduction time (half of I_{pk}), *not* the V_f at the average current.

Now the diode current ramps down from $I_{pk,secondary} = I_{pk,primary} \times$ turns ratio $= 0.956A \times 10 = 9.56A$ to zero. The forward voltage in the schottky during its conduction time, following the approximation just mentioned, is given by the V_f at half its peak current of $(9.56A/2) = 4.8A$, for which, from Figure 8.3, $V_f = 0.32V$; and so the power in the schottky is approximately $P = I_{avg} \times V_f = 2.0A \times 0.32V = 640mW$. Note that this is by far the largest loss in the converter—the power lost in the diode is 50% more than that in the transistor! This is the problem with the discontinuous flyback topology—the currents are very high even for very moderate power levels. You should observe, though (as a slight mitigating factor) that for the discontinuous flyback, there is no reverse recovery loss in the diode, even if it weren't a schottky, because the diode current goes to zero before a reverse voltage is applied, as long as the reverse recovery time is fast enough (i.e., faster than the converter off-time, [4μs − transistor on-time − diode on-time]). Other topologies' losses may, however, strongly depend on diode reverse recovery time.

Electrical Characteristics @ T_J = 25°C (unless otherwise specified)

	Parameter	Min.	Typ.	Max.	Units	Test Conditions
$V_{(BR)DSS}$	Drain-to-Source Breakdown Voltage	200	—	—	V	V_{GS}=0V, I_D= 250μA
$\Delta V_{(BR)DSS}/\Delta T_J$	Breakdown Voltage Temp. Coefficient	—	0.29	—	V/°C	Reference to 25°C, I_D= 1mA
$R_{DS(on)}$	Static Drain-to-Source On-Resistance	—	—	0.80	Ω	V_{GS}=10V, I_D=3.1A ④
$V_{GS(th)}$	Gate Threshold Voltage	2.0	—	4.0	V	V_{DS}=V_{GS}, I_D= 250μA
g_{fs}	Forward Transconductance	1.5	—	—	S	V_{DS}=50V, I_D=3.1A ④
I_{DSS}	Drain-to-Source Leakage Current	—	—	25	μA	V_{DS}=200V, V_{GS}=0V
		—	—	250		V_{DS}=160V, V_{GS}=0V, T_J=125°C
I_{GSS}	Gate-to-Source Forward Leakage	—	—	100	nA	V_{GS}=20V
	Gate-to-Source Reverse Leakage	—	—	-100		V_{GS}=-20V
Q_g	Total Gate Charge	—	—	14	nC	I_D=4.8A
Q_{gs}	Gate-to-Source Charge	—	—	3.0		V_{DS}=160V
Q_{gd}	Gate-to-Drain ("Miller") Charge	—	—	7.9		V_{GS}=10V See Fig. 6
$t_{d(on)}$	Turn-On Delay Time	—	7.2	—	ns	V_{DD}=100V
t_r	Rise Time	—	22	—		I_D=4.8A
$t_{d(off)}$	Turn-Off Delay Time	—	19	—		R_G=18Ω
t_f	Fall Time	—	13	—		R_D=20Ω ④
L_D	Internal Drain Inductance	—	4.5	—	nH	Between lead, 6 mm (0.25in.) from package and center of die contact
L_S	Internal Source Inductance	—	7.5	—		
C_{iss}	Input Capacitance	—	260	—	pF	V_{GS}=0V
C_{oss}	Output Capacitance	—	100	—		V_{DS}=25V
C_{rss}	Reverse Transfer Capacitance	—	30	—		f=1.0MHz See Figure 5

Source-Drain Ratings and Characteristics

	Parameter	Min.	Typ.	Max.	Units	Test Conditions
I_S	Continuous Source Current (Body Diode)	—	—	5.2	A	MOSFET symbol showing the integral reverse p-n junction diode.
I_{SM}	Pulsed Source Current (Body Diode) ①	—	—	18		
V_{SD}	Diode Forward Voltage	—	—	1.8	V	T_J=25°C, I_S=5.2A, V_{GS}=0V ④
t_{rr}	Reverse Recovery Time	—	150	300	ns	T_J=25°C, I_F=4.8A
Q_{rr}	Reverse Recovery Charge	—	0.91	1.8	μC	di/dt=100A/μs ④
t_{on}	Forward Turn-On Time	Intrinsic turn-on time is negligible (turn-on is dominated by L_S+L_D)				

Notes:

① Repetitive rating; pulse width limited by max. junction temperature

② V_{DD}=50V, starting T_J=25°C, L=6.1mH R_G=25Ω, I_{AS}=5.2A

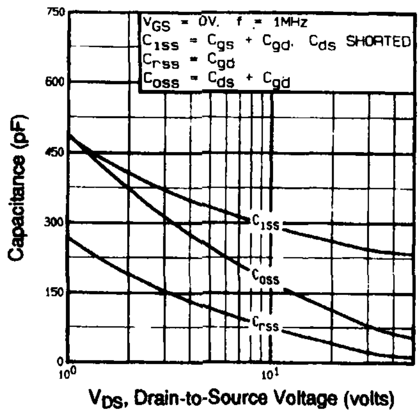
③ I_{SD}≤5.2A, di/dt≤95A/μs, V_{DD}≤$V_{(BR)DSS}$, T_J≤150°C

④ Pulse width ≤ 300 μs; duty cycle ≤2%.

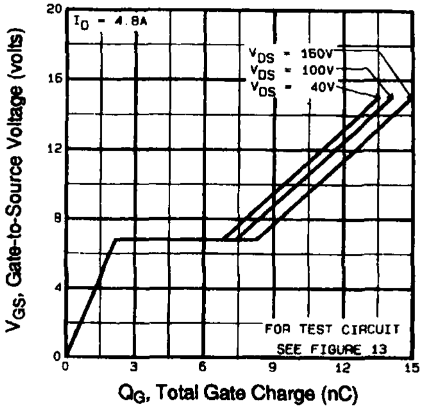
Figure 8.2 Data sheet for the IRF620. (From Ref. 1.)

The final loss to be considered in the power stage is that due to the ripple current in the output capacitor heating the ESR (for the sake of simplicity, this example calculation ignores losses in the input capacitor, EMI filter, fuse, etc.; these can be included by multiplying their RMS current squared times their resistance). In most converters, the loss in the output capacitor is negligible because an inductor smoothes the current before it gets to the cap. In a discontinuous mode flyback, however, we have a real mess: the peak currents are very high and go directly into the capacitor, and thus the losses can be substantial. The complete calculation is gone through in this example in order to provide a guideline for future calculations by the reader—because it is quite a mess!

First we have to know how long the diode is conducting. To calculate this, we must find how long the current in the primary of the transformer takes to return to zero (since



Typical Capacitance Vs.
Drain-to-Source Voltage



Typical Gate Charge Vs.
Gate-to-Source Voltage

Figure 8.2 (*Continued*)

this current is reflected from the secondary current through the diode), which in turn is determined by the output voltage reflected back to the primary inductance by the turns ratio. In detail, the 5V output is actually about 5.3V on the anode side of the schottky. This reflects back to the primary as 53V in addition to the DC input (10 : 1 transformer). Since the voltage is across a transformer, it is AC; that is, the total voltage on the drain is not 53V, but rather $V_{DS} = \text{AC} + \text{DC} = 53\text{V} + 48\text{V} = 101\text{V}$. Thus the (reflected) inductor current ramps down at $53\text{V}/93\mu\text{H} = 570\text{mA}/\mu\text{s}$. Starting from the peak primary current of 956mA, it then takes $t = 956\text{mA}/(570\text{mA}/\mu\text{s}) = 1.677\mu\text{s}$ for the current to return to zero. (What we have really done is to reset the flux in the core.) Incidentally, this tells us that the converter is indeed operating in discontinuous conduction mode: $1.677\mu\text{s} = 0.419$ duty cycle, and the on-time of the transistor is 0.463, so both transistor and diode are nonconducting for $1 - (0.419 + 0.463) = 0.12$ of the period.

This rather roundabout method for calculating the diode conduction time is in fact the only way.

To calculate the losses in the capacitor, it must be remembered that only AC current goes through it. We already know the current in the diode, and the output current is 2ADC, so the capacitor current can be determined as shown in Figures 8.4 and 8.5, where current into the capacitor is defined to be positive. When the diode is conducting 9.56A, for example, 2A goes out to the load, the rest goes into the capacitor; when the diode is nonconducting, the 2A load comes from the capacitor. Since there is always 2A flowing out to the load, the 9.56A peak in the diode contributes only 7.56A into the capacitor, and progressively less as the diode current ramps down. We have already determined that the time when the diode current goes to zero corresponds to a duty cycle of 0.42. The capacitor current goes to zero at a time determined again by the reflection back to primary, as shown in Figure 8.5 and expressed in the following equation:

$$t = L\frac{I}{V} = 93\mu\text{H}\frac{7.56\text{A}/10}{53\text{V}} = 1.33\mu\text{s} = 0.33 \text{ of the period}$$

MOTOROLA
SEMICONDUCTOR
TECHNICAL DATA

1N5823, 1N5824 1N5825

1N5823 and 1N5825 are Motorola Preferred Devices

Designer's Data Sheet

Power Rectifiers

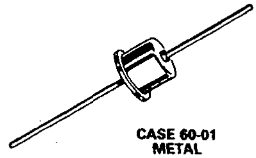
. . . employing the Schottky Barrier principle in a large area metal-to-silicon power diode. State-of-the-art geometry features chrome barrier metal, epitaxial construction with oxide passivation and metal overlap contact. Ideally suited for use as rectifiers in low-voltage, high-frequency inverters, free-wheeling diodes, and polarity-protection diodes.

- Extremely Low v_F
- Low Power Loss/High Efficiency
- Low Stored Charge, Majority Carrier Conduction

Mechanical Characteristics:
- Case: Welded steel, hermetically sealed
- Weight: 2.4 grams (approximately)
- Finish: All External Surfaces Corrosion Resistant and Terminal Leads are Readily Solderable
- Polarity: Cathode to Case
- Shipped 50 units per tray
- Marking: 1N5823, 1N5824, 1N5825

SCHOTTKY BARRIER RECTIFIERS

5 AMPERE
20, 30, 40 VOLTS



CASE 60-01
METAL

*MAXIMUM RATINGS

Rating	Symbol	1N5823	1N5824	1N5825	Unit
Peak Repetitive Reverse Voltage Working Peak Reverse Voltage DC Blocking Voltage	V_{RRM} V_{RWM} V_R	20	30	40	Volts
Non-Repetitive Peak Reverse Voltage	V_{RSM}	24	36	48	Volts
RMS Reverse Voltage	$V_{R(RMS)}$	14	21	28	Volts
Average Rectified Forward Current $V_{R(equiv)} \leq 0.2\ V_{R\ (dc)}$, $T_C = 75°C$ $V_{R(equiv)} \leq 0.2\ V_{R\ (dc)}$, $T_L = 80°C$ $R_{\theta JA} = 25°C/W$, P.C. Board Mounting,	I_O	← 15 → ← 5.0 →			Amp
Ambient Temperature Rated $V_{R\ (dc)}$, $P_{F(AV)} = 0$ $R_{\theta JA} = 25°C/W$	T_A	65	60	55	°C
Non-Repetitive Peak Surge Current (Surge applied at rated load conditions, halfwave, single phase 60 Hz)	I_{FSM}	← 500 (for 1 cycle) →			Amp
Operating and Storage Junction Temperature Range (Reverse Voltage applied)	T_J, T_{stg}	← −65 to +125 →			°C
Peak Operating Junction Temperature (Forward Current Applied)	$T_{J(pk)}$	← 150 →			°C

*THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	3.0	°C/W

*ELECTRICAL CHARACTERISTICS (T_C = 25°C unless otherwise noted)

Characteristic	Symbol	1N5823	1N5824	1N5825	Unit
Maximum Instantaneous Forward Voltage (1) (i_F = 3.0 Amp) (i_F = 5.0 Amp) (i_F = 15.7 Amp)	v_F	 0.330 0.360 0.470	 0.340 0.370 0.490	 0.350 0.380 0.520	Volts
Maximum Instantaneous Reverse Current @ rated dc Voltage T_C = 25°C T_C = 100°C	i_R	 10 100	 10 125	 10 150	mA

(1) Pulse Test: Pulse Width = 300 μs, Duty Cycle = 2.0% *Indicates JEDEC Registered Data for 1N5823–1N5825

Figure 8.3 Data sheet for the 1N5825. (From Ref. 2.) (Copyright of Motorola, used by permission.)

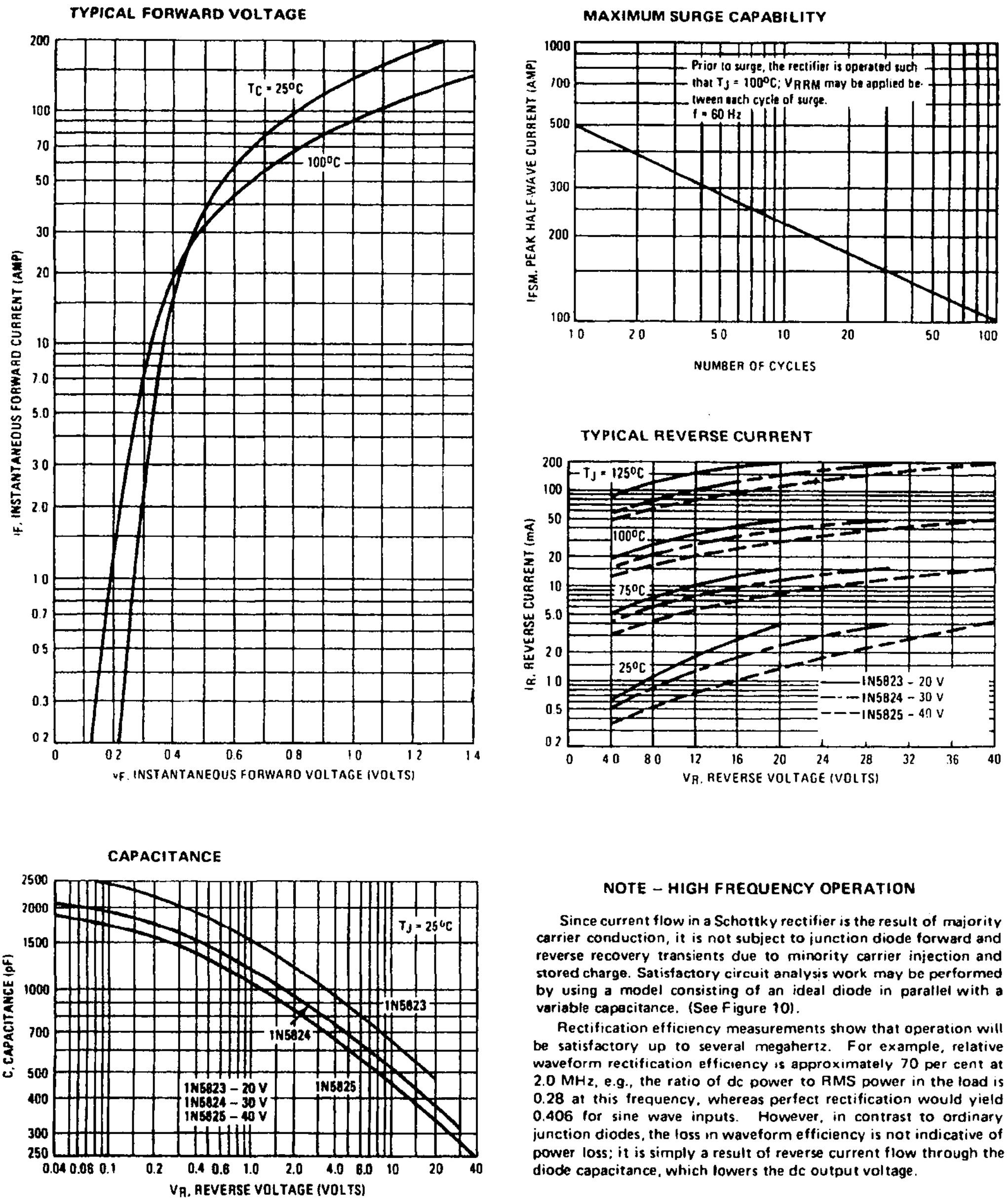


Figure 8.3 *(Continued)*

As a check, we can verify that the average capacitor current is zero:

$$I_{avg} = \frac{7.56\text{A}}{2} \times 0.33 + \frac{-2\text{A}}{2} \times (0.42 - 0.33) + (-2\text{A}) \times (1 - 0.42) = 0$$

Now we are ready to calculate the RMS current. Remember that RMS is calculated by first squaring the current (RMS = root-mean-*square*), so that the AC current in Figure 8.6 is conceptually equivalent. This is true because squaring a negative is equal to squaring

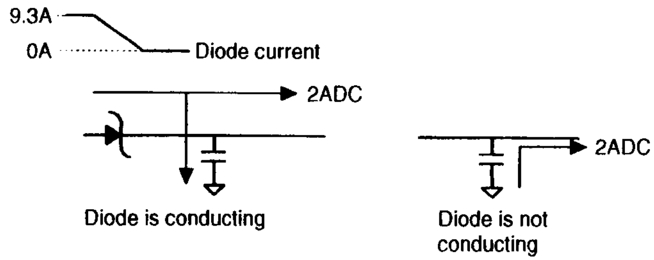


Figure 8.4 Currents in the secondary of a discontinuous mode flyback.

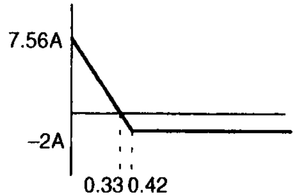


Figure 8.5 Capacitor currents are found by subtracting the load current from the diode current.

its absolute value. From this it is now straightforward to calculate the RMS current: it is the RMS of the individual RMS current pieces of this waveform. The first piece is

$$I_1 = 7.56\sqrt{\frac{0.33}{3}} = 2.51\text{A}$$

the second piece is

$$I_2 = 2.00\sqrt{\frac{0.42-0.33}{3}} = 0.35\text{A}$$

and the third piece is

$$I_3 = 2.00\sqrt{1-0.42} = 1.52\text{A}$$

so that the total RMS current is

$$I_{\text{rms}} = \sqrt{I_1^2 + I_2^2 + I_3^2} = 2.96\text{A}$$

Now at last we can calculate the losses. Suppose the output capacitor to consist of two paralleled 20μF, 10mΩ ESR MLC capacitors. (These are selected because 7.5A peak capacitor current times 5mΩ gives 38mV ripple, which is reasonable for a 5V output.) Losses in the caps are thus $(2.96\text{A})^2 \times 5\text{m}\Omega = 43\text{mW}$, or 21mW per package. Note that if

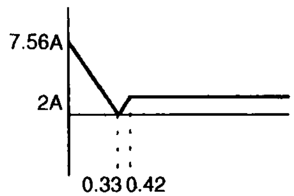


Figure 8.6 Equivalent currents for calculating RMS current.

a large aluminum had been chosen instead, with an ESR, say, of 25mΩ, not only would the ripple be up to 190mV, but the losses would be 215mW, possibly overheating it.

As a final source of loss in the converter, we suppose that the PWM is running from a 10V winding on the flyback (not shown in Figure 8.1), and that it draws 10mA of current (in addition to the gate current which has already been accounted for). If the rectifier on the winding drops 1V, the PWM uses (10V + 1V)10mA = 110mW.

We can now calculate the efficiency using Table 8.1.

TABLE 8.1 Losses for Example 1 Converter

Item	Losses (mW)
Magnetics	150
MOSFET $R_{DS,on}$	141
MOSFET switching	287
MOSFET gate charge	22
Diode	640
Capacitor	43
IC	110
Total	1393

Output power is 10W, so input power is 10W plus the losses, or

$$\eta = \frac{10\mathrm{W}}{11.393\mathrm{W}} = 88\%$$

which is quite good for this type of converter at this power level.

Before leaving this example, let's note a small inconsistency. We initially assumed that the power in the converter was 10.6W, accounting for the output power and the approximate losses in the rectifier. In the end, however, the actual power level was 11.4W, which is 800mW higher. To be fully consistent, we could now go back and redo the calculations at this higher power level. However, it is easy to see that the overall efficiency is going to remain close to 88%; since the difference was only 800mW, the additional losses calculated this second time through would be approximately 800mW × (1 − 90%) = $80mW$, which can be ignored for the level of approximation being used in this example.

Example Calculation 2

Let's now look at the same converter, but producing only 1W (5V at 200mA) instead of 10W. To make life easy, we'll assume that the magnetic core has been resized, but the inductance has remained the same, changing only the duty cycle appropriately, and that the magnetics' total losses are the same; all other items on the bill of materials will remain the same, as well. Repeating all the calculations yields the following results:

$$\eta = \frac{1\mathrm{W}}{1.422\mathrm{W}} = 70\%$$

The most significant thing to notice here (aside from the overall drop in efficiency, which always happens at lighter loads) is that the distribution of losses has changed (see

Table 8.2). Whereas in the first example the MOSFET losses were about two-thirds those of the diode (450mW < 640mW), so that it would make sense to consider synchronous rectification, in the second example, the MOSFET losses are double those of the diode's (116mW > 44mW), and so synchronous rectification would actually make the efficiency *worse*.

TABLE 8.2 Losses for Example 2 Converter

Item	Losses (mW)
Magnetics	150
MOSFET $R_{DS,on}$	4
MOSFET switching	90
MOSFET gate charge	22
Diode	44
Capacitor	2
IC	110
Total	422

Practical Note Rather than tediously calculating tables of losses, repeatedly trying to find an optimum, you can enter the various formulas into a spreadsheet, allowing easy parametric variation.

Improving Efficiency

From the two sample calculations above, it should be clear that achieving efficiency improvements is rather different for low power converters versus those in the medium and high ranges. For the latter cases, the most important way to increase efficiency is to lower the MOSFETs switching frequency, because this reduces the switching loss. This *does* make all the components physically larger, which is the usual trade-off. A second way of increasing the efficiency of medium and high power converters is to use synchronous rectification. Of course this is difficult if the supply is isolated, but whenever practical, replacing the diode, even a schottky, with a MOSFET will substantially reduce the loss, as is obvious from the first example. The underlying problem with the design in the first example is that it is a discontinuous mode flyback, so that the currents are high, generating high losses in both the transistor and the diode (and in the capacitor, had it not had very low ESR); selecting a different topology would generate substantially lower losses, though at the cost of additional components. It should also be clear that to maximize efficiency, extraneous losses should be avoided. That is, preloads either should not be used at all or turned off when not needed, the start-up circuit should be turned off after the converter is started, and so on.

For the case of a low power converter, efficiency improvements are a little more difficult to come by. Reducing the switching frequency is still number one in importance, although now the savings may be due more to reductions in gate charge loss than in switching loss. Contrary to the case of higher power converters, synchronous rectification at low power may give *lower* efficiency than a diode. To repeat, this is because the losses due to increased gate charge losses may be greater than the savings due to the decreased conduction loss. To overcome this limitation, certain modern PWM ICs actually reduce their switching frequency when a light load is detected. Others go into "pulse frequency modulation" (PFM) in which the transistor is turned on only when the voltage has drooped to a certain level. Yet others turn off synchronous rectification at low power, running only on the paralleled schottky.

A final tactic that may yield efficiency improvement in low power converters is reducing IC current: some ICs take as much as 30mA of current to run, and at low power this can be a substantial hit on efficiency. In the second example, the loss due to IC current was one-fourth of the total losses!

THERMAL MANAGEMENT

Thermal management is an important part of many converter designs, and not only for the obvious reason that things burn up if they get too hot. The effect of temperature on component life, and thus power supply life, is explored immediately below. Besides, in consumer applications it may be undesirable to have a supply that is so hot that consumers can burn their fingers by touching it! Even if thermal management per se isn't necessary, your efficiency calculations have depended on knowing the temperature of the various components, and so being able to calculate temperatures is important to assuring that the desired efficiency is indeed achieved.

Component Life versus Temperature

The life expectancy of every component in a power supply depends on its temperature: if the temperature rises, the life expectancy decreases. This relationship directly affects the field failure rate of your converter, especially if any of the components are being run close to their maximum rated temperature.

Practical Note As a rule of thumb, the life of a component approximately doubles for each 20°C drop in temperature. Thus, a capacitor rated at 2000 hours at 105°C will have a life of approximately

$$2000\text{h} \times 2^{(105°\text{C}-25°\text{C})/20°\text{C}} = 32{,}000\text{h} \approx 4 \text{ years}$$

at 25°C.

The most obvious example of this temperature dependence of operational life, and the most consistently and damagingly overlooked in power supply design, is the case of aluminum electrolytic capacitors. As mentioned in Chapter 3, and as the example in the Practical Note suggests, aluminums have very low life expectancy at their rated temperature, which rating is often 105°C or even 85°C—remember that 2000 hours is less than 3 months.

Practical Note For most power supply applications that call for aluminum electrolytics, you will be using 105°C parts; in most cases these parts should be rated 2000 hours or better yet 5000 hours. Give some serious thought to using tantalums instead!

As a mitigating factor, you should realize that supplies won't be operated 24 hours a day at their maximum temperature. If you can estimate the percentages of time spent at different temperatures, you will get a far more favorable estimate of the capacitor's life.

How does an aluminum electrolytic manifest that it has reached its end of life? The author ran a power supply using aluminums at high temperature for a year. Over that year, the ESR of the capacitors increased, at first slowly, and then rapidly. At the end of the year, the ESR was so high that the output ripple voltage of the supply was wildly beyond spec. Thus, running a capacitor beyond its rated life can be expected to result in a power supply failing spec, and possibly damaging the components it's supposed to be powering.

Another area of potential concern is the temperature rating of ICs. There are three temperature grades of ICs: commercial, which is rated from 0°C to 70°C; industrial, rated from −40°C to 85°C; and military, rated from −55°C to 125°C. Now of course the manufacturers of the parts don't make different dice for the differing temperature grades; the difference is in the packaging (plastic for commercial and industrial, ceramic for military) and the temperature over which the components are tested, that is, over which their operation is guaranteed. So operating a commercial part at 90°C probably won't cause any operational problem. Your worst-case analysis will be problematic, however, the MTBF will be bad (as indicated above), and if the part *does* fail, the manufacturer will be justified in claiming no responsibility.

A final topic deserving a mention is the temperature of MOSFETs. In the calculation of efficiency done earlier in this chapter, it was assumed that the MOSFET reached a stable operating temperature of 60°C, and the losses were calculated based on this temperature. However, it should be noted that the $R_{DS,on}$ of a MOSFET depends on its temperature, so that losses are temperature dependent, and of course temperature depends on losses. Thus a MOSFET can produce enough heat to make its temperature rise, which causes the resistance to rise, which causes losses to rise, which soon causes the device to exceed its rated temperature. The end result of such thermal runaway, of course, is failure.

Given all this, it is certainly highly desirable to use parts that are rated for the temperature they're going to see. On the other hand, there is a price differential between the temperature grades—moderate for a change from commercial to industrial temperatures, but extremely steep for industrial to military. Thus, holding down the overall temperature in the converter is imperative, not only for maintaining converter life, but also if cost is at all important.

Modules

Talking about component temperatures, we once again come to converter modules. The same reasons that drive manufacturers to quote unrealistic efficiencies result in the giving out of unrealistic estimates of the amount of output power the modules can produce. The limiting factor in output power is the amount of heat generated inside the module: the two are of course proportional. The problem is that if you just solder a module to a PC board and try to draw the rated power out, the module will burn up. Closer inspection of the module's data sheet reveals that the rated power is available only if there is attached to the module, *a heat sink that is larger than the module!* So the nice low profile power supply has suddenly doubled in height, or else you must buy a module that is (apparently) very overrated for the application, and therefore much more expensive.

MIL-HDBK-217

After all these dire warnings about the effects of temperature on converter life, how about a method for calculating MTBF, to see whether your design is going to meet its specified life? One standard way is to use MIL-HDBK-217 [3]. The U.S. military maintains an ongoing program on the failure rates of many common components, and the information gleaned is put into a useful form in this handbook, which is updated periodically (thus the F in MIL-HDBK-217F indicates the publication's sixth revision). We'll first do a quick sample calculation, and then discuss some of the possible concerns associated with using 217.

MIL-HDBK-217: Example

To give a simple example of the usage of MIL-HDBK-217, let's suppose that we are trying to establish the MTBF of a system of three paralleled aluminum electrolytic capacitors. The table of contents of 217F [3] shows two sections covering aluminums; one covers "non-established reliability" parts (i.e., commercial parts), and so we use this [3, pp. 10-24 to 10-25].

Examination of Figure 8.7 reveals that λ_p (in failures per million hours), of an aluminum electrolytic capacitor is the product of four factors. The first factor, λ_b, is the base failure rate, and it depends on the temperature rating of the capacitor. Let's suppose that this is 105°C, so we use the table for λ_b ($T = 105$°C Max Rated) in Figure 8.7. Suppose the average temperature the capacitor is going to see during its life is 60°C. (Again, it is important to use *average* temperature, not maximum.) Our reference, 217F, also needs the "stress" on the capacitor, which it defines as the ratio of operating to rated voltage: suppose that the capacitor is rated at 5V, and we are applying in steady state 3.5V, so the stress $S = 0.7$. (Again, be sure to use *average* voltage, not maximum.) We then find that $\lambda_b = 0.14$.

The next factor is π_{CV}, the capacitance factor. Let's suppose each capacitor to be 1000µF. Now, 1000µF isn't listed in the table, so we can use the formula instead:

$$\pi_{CV} = 0.34C^{0.18} = 0.34(1000)^{0.18} = 1.18 \approx 1.2$$

Since all the factors in the table for π_{CV} are rounded off to two significant digits, the implication is that this is the accuracy of the formula.

The third factor, π_Q, is easy: this is a commercial capacitor, and so it is the lowest possible quality factor, 10.

Finally, the fourth factor, π_E, is for the environment. All commercial supplies operate in "ground, benign" conditions, and so $\pi_E = G_B = 1.0$.

We now find that the failure rate for a single capacitor under these conditions is

$$\lambda_p = \lambda_b \pi_{CV} \pi_Q \pi_E = 0.14 \times 1.2 \times 10 \times 1.0 = 1.68$$

MIL-HDBK-217F

10.14 CAPACITORS, FIXED, ELECTROLYTIC, ALUMINUM

SPECIFICATION	STYLE	DESCRIPTION
MIL-C-39018	CUR and CU	Electrolytic, Aluminum Oxide, Est. Rel. and Non-Est. Rel.

$$\lambda_p = \lambda_b \pi_{CV} \pi_Q \pi_E \text{ Failures}/10^6 \text{ Hours}$$

Base Failure Rate - λ_b
(T = 85°C Max Rated)
(MIL-C-39018 Style 71)

T_A (°C)	Stress .1	.3	.5	.7	.9
0	.0095	.011	.019	.035	.064
10	.012	.015	.024	.046	.084
20	.017	.020	.033	.062	.11
30	.023	.028	.046	.087	.16
40	.034	.042	.068	.13	.23
50	.054	.065	.11	.20	.36
60	.089	.11	.18	.33	.60
70	.16	.19	.31	.58	1.1
80	.29	.35	.58	1.1	2.0

$$\lambda_b = .00254\left[\left(\frac{S}{.5}\right)^3 + 1\right]\exp\left(5.09\left(\frac{T+273}{358}\right)^5\right)$$

T = Ambient Temperature (°C)

S = Ratio of Operating to Rated Voltage

Operating voltage is the sum of applied D.C. voltage and peak A.C. voltage.

Base Failure Rate - λ_b
(T = 105°C Max Rated)
(MIL-C-39018 Styles 16 and 17)

T_A (°C)	Stress .1	.3	.5	.7	.9
0	.0070	.0084	.014	.026	.047
10	.0085	.010	.017	.031	.057
20	.011	.013	.021	.040	.072
30	.014	.017	.027	.051	.094
40	.019	.022	.037	.069	.13
50	.026	.031	.052	.097	.18
60	.038	.046	.076	.14	.26
70	.059	.071	.12	.22	.40
80	.095	.11	.19	.35	.64
90	.16	.20	.32	.61	1.1
100	.30	.36	.59	1.1	2.0

$$\lambda_b = .00254\left[\left(\frac{S}{.5}\right)^3 + 1\right]\exp\left(5.09\left(\frac{T+273}{378}\right)^5\right)$$

T = Ambient Temperature (°C)

S = Ratio of Operating to Rated Voltage

Operating voltage is the sum of applied D.C. voltage and peak A.C. voltage.

Base Failure Rate - λ_b
(T = 125°C Max Rated)
(All MIL-C-39018 Styles Except 71, 16 and 17)

T_A (°C)	Stress .1	.3	.5	.7	.9
0	.0055	.0067	.011	.021	.038
10	.0065	.0078	.013	.024	.044
20	.0077	.0093	.015	.029	.052
30	.0094	.011	.019	.035	.064
40	.012	.014	.023	.044	.080
50	.015	.019	.030	.057	.10
60	.021	.025	.041	.077	.14
70	.029	.035	.057	.11	.20
80	.042	.050	.083	.16	.28
90	.064	.077	.13	.24	.43
100	.10	.12	.20	.38	
110	.17	.21	.34	.63	
120	.30	.37	.60	1.1	

$$\lambda_b = .00254\left[\left(\frac{S}{.5}\right)^3 + 1\right]\exp\left(5.09\left(\frac{T+273}{398}\right)^5\right)$$

T = Ambient Temperature (°C)

S = Ratio of Operating to Rated Voltage

Operating voltage is the sum of applied D.C voltage and peak A.C. voltage.

Figure 8.7 MIL-HDBK-217F reliability data for aluminum electrolytics.

failures per million hours (or 1680 FITs: one FIT = one failure per billion hours). As such things go, this is pretty high, showing that aluminums are poor components; many components will have only some tens of FITs. The predicted MTBF for this capacitor is

$$\text{MTBF} \equiv \frac{1}{\lambda} = \frac{1{,}000{,}000\text{h}}{1.68} = 600{,}000\text{h}$$

MIL-HDBK-217F

10.14 CAPACITORS, FIXED, ELECTROLYTIC, ALUMINUM

Capacitance Factor - π_{CV}

Capacitance, C (µF)	π_{CV}
2.5	.40
55	.70
400	1.0
1700	1.3
5500	1.6
14,000	1.9
32,000	2.2
65,000	2.5
120,000	2.8
$\pi_{CV} = .34C^{0.18}$	

Quality Factor - π_Q

Quality	π_Q
S	.030
R	.10
P	.30
M	1.0
Non-Est. Rel.	3.0
Lower	10

Environment Factor - π_E

Environment	π_E
G_B	1.0
G_F	2.0
G_M	12
N_S	6.0
N_U	17
A_{IC}	10
A_{IF}	12
A_{UC}	28
A_{UF}	35
A_{RW}	27
S_F	.50
M_F	14
M_L	38
C_L	690

Figure 8.7 (*Continued*)

In our example, we have three of these capacitors in parallel. The total failure rate is the sum of the failure rates for each component (in a simple model) and so the total failure rate is 5040 FITs, for an MTBF of 200,000 hours.

What (besides using some other style of capacitor) can be done to improve this MTBF? The biggest factor in this case is voltage rating. In agreement with the rule of thumb given above, dropping the temperature from 60°C to 40°C reduces λ_b by a factor of 2, from 0.14 to 0.069. As a practical matter, however, it is much easier to go to a higher voltage capacitor: using a 10V cap reduces the stress from 0.7 to 0.35, reducing λ_b from 0.14 to 0.051, almost a factor of 3!

MIL-HDBK-217: Discussion

You should be aware of a potential problem in the use of MIL-HDBK-217. Since the handbook is designed for use on designs of military equipment, commercial parts are not always covered in the depth you could wish—practically, you sometimes have to guess which of the possible choices most closely matches the part you are actually using, as in the example above.

You sometimes hear people argue that the MTBFs derived from 217 are too pessimistic. These people sometimes cite a Bellcore reliability manual as giving far longer lifetimes. The author's experience is that 217 gives quite realistic estimates. Just be careful to verify that when a converter is advertised with an MTBF of such-and-such, the manufacturer used 217, not something else (such as imagination), and that the calculation was done using actual stresses, not the "parts count" method, which is based on a count of the number of components of a certain type that are likely to be used in a design. Parts count is supposed to be for preliminary estimates of reliability only, not for calculating MTBF of a finished design.

In line with the cautions elsewhere in this book, you might also want to be careful about the use of programs that calculate MTBF from 217 for you. This type of software can certainly save some drudgery, but how certain are you that the programmers typed in all the formulas correctly? Before using this software routinely, you would be well advised to check one calculation by hand for each type of component.

Temperature Calculation

After all this discussion of temperature, it's time to calculate the temperature of an actual component. Given a component's power dissipation and its thermal interfaces, this task is straightforward. It turns out there is an exact analogy between thermal and electrical characteristics, as shown in Table 8.3. (Mechanical engineers use many other units, also; the best plan for electrical engineers is to avoid confusion by converting these other units to the units shown here.) This analogy directly implies that if you have two thermal interfaces in series, their thermal resistances add.

EXAMPLE

The IRF620 used in the efficiency calculation in Example Calculation 1 was dissipating 450mW at 60°C. From its data sheet (Figure 8.2), we see that it has a thermal resistance from junction to case (i.e., from the actual die to the outside of the TO-220 package) of $\Theta_{JC} = 2.5°C/W$, and a thermal resistance from the case to sink (i.e., from the TO-220 package, through thermal compound, to a heat sink) of 0.5°C/W. Let's suppose the heat sink has a further thermal resistance of about 40°C/W to the point where the temperature is held fixed at 45°C. The electrical analogy is a current source of magnitude 450mA, attached through three series resistors, of resistance 2.5, 0.5, and 40Ω, respectively, to a voltage source of 45V. Clearly, the top of the resistor stack is $45V + 450mA(2.5\Omega + 0.5\Omega + 40\Omega) = 64V$, or 64°C. In this example, the die is only a degree hotter than the case, but it does not always work out this way.

TABLE 8.3 Correspondence Between Thermal and Electrical Characteristics

Thermal	Units	Corresponds to	Electrical	Units
Temperature	°C		Voltage	volts
Heat source	watts		Current	amps
Thermal resistance	°C/Watt		Resistance	ohms
Thermal capacitance	joules/°C		Capacitance	farads
Thermal time constant	seconds		*RC* time constant	seconds

The analogy between thermal characteristics and electrical circuits also extends to thermal capacitance, although it is more usual to find the thermal time constant specified, or just a graph of thermal impedance as a function of time to be shown.

EXAMPLE

Let's suppose that the IRF620 was dissipating 10W. Clearly, the die temperature would rise so far that the device would fail, because $45V + 10A(2.5\Omega + 0.5\Omega + 40\Omega) = 475V$ or 475°C! However, suppose instead that the 10W were dissipated for only 100μs, after which the power dissipation returned to 450mW. The curve for thermal response (Figure 8.2) indicates that a 100μs single pulse has a thermal impedance that is one-tenth its steady-state response; we'll suppose that the rest of the system has the same equivalent thermal time constant. Thus, at the end of this time, the temperature has risen to $64V + [10A(2.5\Omega + 0.5\Omega + 40\Omega)0.1] = 107V$ or 107°C, which is entirely tolerable for the device. We can also find the thermal capacitance from this information, if we need to: since the thermal resistance is 2.5°C/W, the thermal capacitance must be $C = t/R = 100\mu s/(2.5°C/W) = 40\mu J/°C$.

It is for exactly this same reason, namely thermal capacitance, that the pulse power in a wirewound resistor can be much higher than its steady-state power dissipation, as discussed in Chapter 3.

Heat Sinks, etc.

The traditional method for getting extra heat out of a device (besides convection; radiation is usually negligible) is a heat sink (i.e., conduction). The heat sink provides a path in addition to convection for the heat, analogous to providing a second resistor in parallel with the first: since this reduces the total resistance, the temperature rise is similarly reduced.

The cheapest type of heat sink is just a piece of metal, frequently anodized, attached to the device being heat-sunk by a clip or screw selected from the huge variety of shapes and sizes available. (Screws are thermally better than clips, because clips have considerable variation in the amount of pressure they provide; but of course screw attachment requires an additional component and additional labor.) Because the device being heat-sunk and/or the heat sink may not be perfectly flat, it is common to apply thermal compound (grease) to the back of the device before attaching it to the heat sink. This fills in the voids, lowering the total thermal resistance. However, thermal compound is quite a mess, getting all over everything, and so its use in production lines tends to be frowned on.

> **Practical Note** If you turn too hard on the screw attaching the component to the heat sink, the metal may actually bow (form a shape like an arch), which leaves an air gap between the device and the heat sink, defeating the heat sink's purpose. Component packages usually specify the maximum torque to be applied, and special screwdrivers are available with calibrated torque, to stop turning the screw when the specified torque is reached.

The metallic heat sink may need to be electrically isolated from the circuit, since the heat sink may be connected to ground, for example, to the enclosure. The isolation can be accomplished by placing an insulator between the component and the heat sink. The most common insulator is a rubbery material called Silpad. Another possibility is the use of mica or beryllium oxide; the latter is often avoided, though, because of concerns about exposure to toxic beryllium dust if the heat sink were to become powdered. (As long as you don't grind the BeO, it will be just fine.)

Heat sinks can work well for applications such as a TO-220 package. However, there doesn't seem to be a good way to attach a heat sink to a surface mount component. This is a particular concern when one is using surface mount MOSFETs: the dominant form of heat transfer for these transistors is through their leads, which can seriously limit the usefulness of the devices in higher power converters. The situation can be somewhat improved by running a large trace underneath the body of the package; unfortunately, manufacturers often neglect to specify the thermal resistance from the die to the body of the package.

If the heat problem in a converter cannot be fixed by means of a heat sink, there are more complex possibilities, such as a fan or a heat pipe. Not only are such solutions expensive, however, their effects are often difficult to calculate as well. For example, knowing a fan's airflow doesn't necessarily tell you much about the amount of cooling a given component will see, because the airflow to that component probably will be partly blocked by other components. Generally, such possibilities should be left to a mechanical engineer who has specialized knowledge in this area.

FEA

A final comment can be made about computer programs for evaluating thermal performance of a converter. There are specialized programs for doing this, using FEA, finite element analysis. Essentially, such software (conceptually) breaks up the converter into small pieces and lets them all interact simultaneously to determine the temperature distribution in the supply, much as an electrical simulation would deal with a network of resistors. FEA programs can even do this on a dynamic basis, using thermal capacitances.

In fact, such a program could also be written using the electrical analogy to thermal properties to map the model into a SPICE model (or using SABER's mixed-mode capabilities directly). It might be interesting to investigate the computational (and economic) efficiency of such a routine versus the rather expensive thermal programs typically used.

REFERENCES

1. *HEXFET Power MOSFET Designer's Manual*, publication HDM-3, 1993. International Rectifier, 233 Kansas St., El Segundo, CA 90245.
2. *Rectifier Device Data*, publication DL151/D, rev. 2, 1995. Motorola, P.O. Box 20912, Phoenix AZ, 85036.
3. MIL-HDBK-217F, Notice 1, in *Military Handbook, Reliability Prediction of Electronic Equipment*, 1992. Rome Laboratory, Griffiss Air Force Base, Rome, NY.