Thermal Considerations In Power BiMOS Low Side Drivers (HIP0080, HIP0081, HIP0082, CA3282 and Others)

Author: Wayne Austin

Overview

The Intersil HIP0080, HIP0081, HIP0082 and CA3282 Power BiMOS Low Side Drivers are designed for use in automotive and industrial switching applications over an operating temperature range of -40^{o}C to +125^{o}C. Each circuit has enhanced performance features that include logic level operation, overload protection and diagnostic feedback by serial interface. The focus of this Application Note is directed to the HIP0082 Power BiMOS Quad Low Side Driver but generally applies to all of the MOS Low Side Drivers. The IC interface and switching configurations are noted in the Block Diagrams below. The CA3282, HIP0081 and HIP0082 are provided in a 15 Lead Power Package with a low junction-to-case thermal resistance and a lead form option for vertical or surface mount. The chip is directly mounted on the copper heat sink tab of the IC to provide a maximum θ_{JC} that is less than 3^{o}C/W.

The designer of a power switching circuit must address many thermal related issues as a part of the design process. A key element in this process is an understanding of the temperature related parameters which are discussed here for educational information. The scope of this material includes thermal related effects on the IC resulting from the externally applied power, load conditions and high ambient temperature operation. Solutions for the problem of pulsed energy generated from inductive switching are derived as source information for the user. As such, the following information is intended to provide the user with a practical working knowledge of the thermal issues common to a Power Switch IC, both from a theoretical and practical approach.

To begin, during switching and in the ON-state heat is generated in the drain (junction) of the MOS driver on the surface of the chip. Both transient and steady state conditions must be controlled to avoid operating temperatures in excess of the maximum junction temperature ratings, usually 150°C in plastic encapsulated chips. The temperature developed at the surface of the chip diffuses from the source of excitation and is conducted to the external surface of the IC package where adequate convection and radiation cooling are required to keep the operating junction temperature within ratings.

Referring to Figure 2 as an illustrated example of the 15 Lead Single-in-Line (SIP) Package, the temperature developed at the active junction of the IC is determined by sustained dissipation over time. Heat flow from the junction-to-case follows the path of least thermal resistance. The junction-to-case thermal resistance is the complex sum of all series and parallel thermal paths that the diffusing heat takes as it flows from the chip junction to the external surface area of the IC. By design, the direct path from the die to heat sink tab and external heat sink is the primary conduit path for escaping heat. The external heat sink is cooled by convection and radiation.

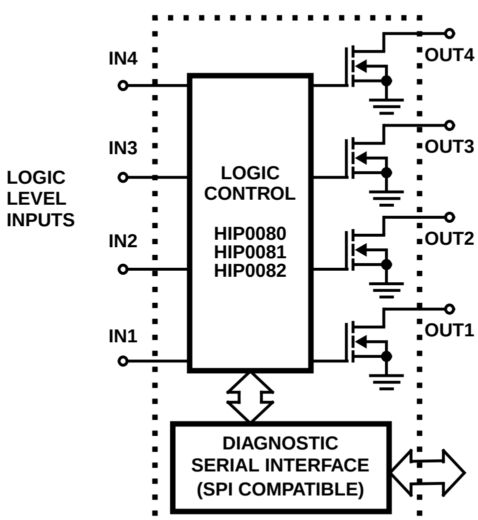

FIGURE 1A.

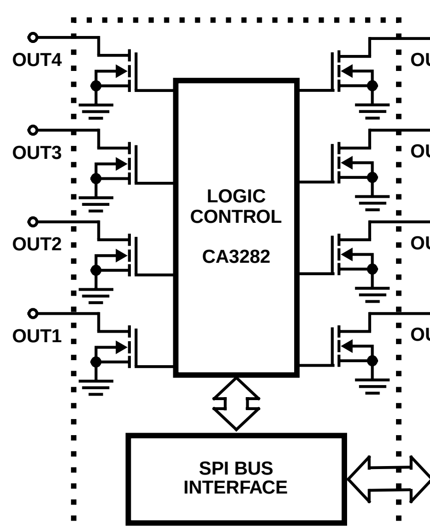

FIGURE 1B.

FIGURE 1. SWITCHING CONFIGURATIONS OF THE QUAD AND OCTAL POWER BiMOS CIRCUITS

To better understand the process of thermal conductivity, background material with thermal definitions and units related to heat are provided in Tables 1 and 2. The SI (International System)[1] symbol definitions provided in Table 1 are limited to the relevant subject material.

Electrical And Thermal Parameters

To understand the process of heat flow, the following parameter definitions are provided:

Temperature, T is the measure of molecular kinetic energy in a body with changes expressed in degrees Celsius or Kelvin. Thermal energy corresponds to temperature difference, ΔT as the potential for heat flow. Thermal energy units are in Btu, gram-calories or kg-calories which, by conversion, are equivalent to energy in Joules.

Specific Heat, c is the heat capacity per unit mass, m_k with units of Joules/kg•K and is often tabulated in calories/gram^{o}C units or Btu/lb•^{o}F in reference publications. Specific Heat is commonly expressed as c_p for constant pressure and c_v for constant volume. Since the difference is small in solids, we may use the general expression,

$$c = \frac{C_T}{m_k} \qquad \text{(EQ. 1)}$$

True specific heat can be a function of temperature and is defined by the differential expression,

$$c(t) = \frac{1}{m_k}\frac{dQ}{dT} \qquad \text{(EQ. 1A)}$$

where Q is heat flow in watts (W).

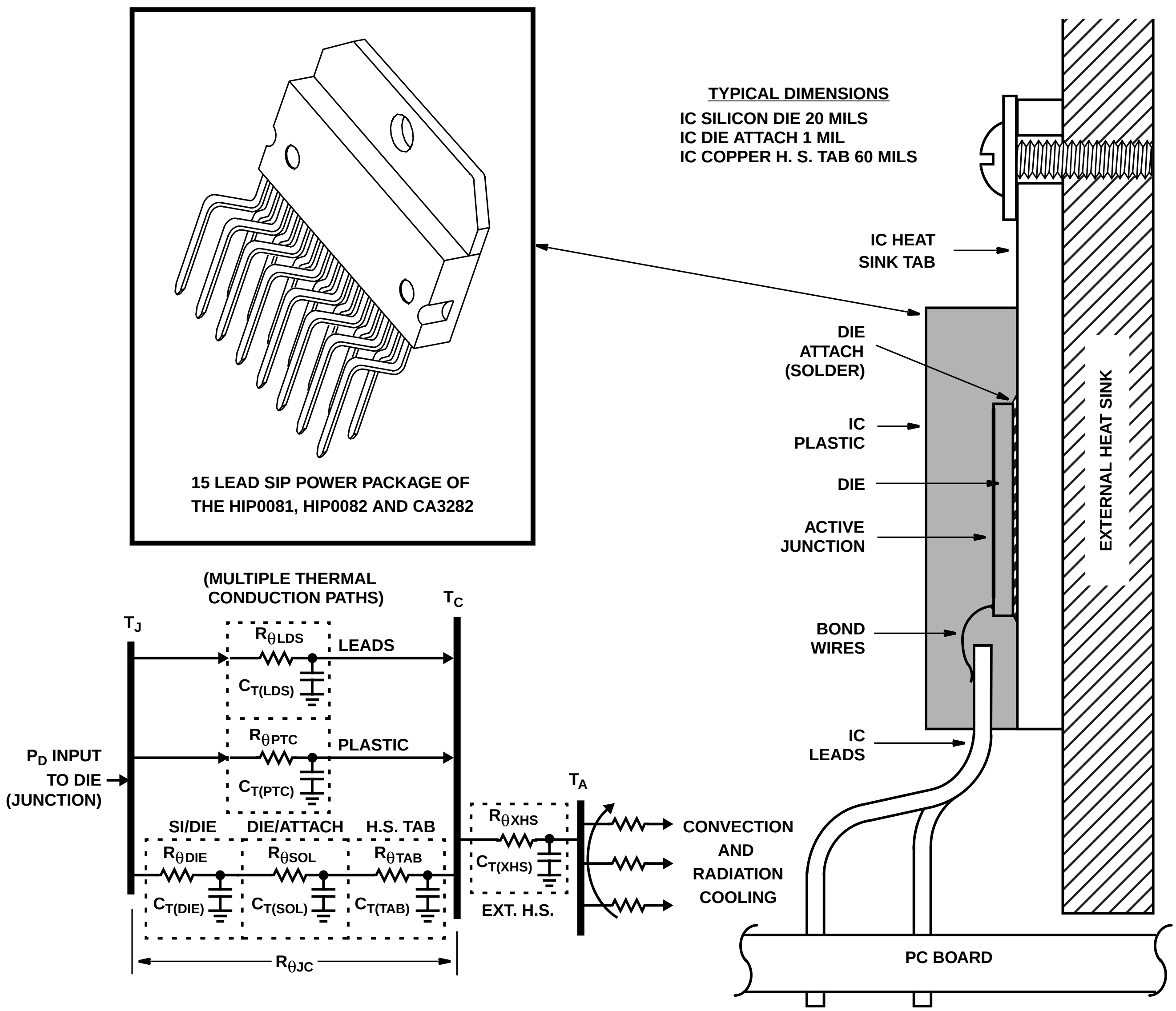

FIGURE 2. TYPICAL HEAT SINK MOUNTING STRUCTURE OF A POWER SWITCHING IC IN THE 15 LEAD SIP POWER PACKAGE. THERMAL RESISTANCE (JUNCTION-TO-AMBIENT) IS PRIMARILY DEFINED BY THE SUM OF ALL THERMAL CON-

TABLE 1. INTERNATIONAL SYMBOLS (SI UNITS)

UNIT	SYMBOL	PARAMETER
ampere	A	Unit of electric current
volt	V	Unit of electromotive force
ohm	Ω	Unit of electric resistance
farad	F	Unit of capacitance
henry	H	Unit of inductance
kilogram	kg	Unit of mass
meter	m	Unit of length
watt	W	Unit of power
joule	J	Unit of energy (heat)
degree	°C	Unit of temperature, Celsius scale
Kelvin	K	Unit of temperature, Kelvin scale
hertz	Hz	Unit of frequency
second	s	Unit of time

TABLE 2. SPECIFIC HEAT AND THERMAL CONDUCTIVITY OF VARIOUS SEMICONDUCTOR MATERIALS

MATERIAL	SPECIFIC HEAT cal/gram•°C	CONDUCTIVITY W/m•°C
Silver	0.056	414
Copper	0.0921	381
Gold	0.031	297
Aluminum	0.226	213
Beryllia	0.286	155
Nickel	0.112	88
Silicon	0.181	See Note 1
Iron	0.108	78
Steel	0.1	46
Lead	0.03	35
5/95 Solder	0.03	35
Alumina	0.18	26
Silver Epoxy	0.15	2.3
Borosilicate Glass	0.199	0.59
Novolac Epoxy	0.2	0.58

NOTE:

1. Silicon Conductivity is temperature dependent. The NIST[6] bibliography reference provides a formula: k = 286/(T-100)(W/cmK) for a range of 300K to 600K. For a nominal range of junction temperatures: k ≈ 400/T (W/cmK) where T is temperature in Kelvin.

Heat Capacity, C_T is the ratio of heat supplied to a composite body for a corresponding temperature rise which leads to the following expression:

$$C_T = \frac{Q}{\Delta T} \qquad \text{(EQ. 2)}$$

Heat Flow, Q is analogous to electrical current and is expressed in joules per second (watt). For a body of mass, m_k Heat Flow may be expressed as:

$$Q = m_k c(T_2 - T_1) \quad \text{or } Q = C_T(T_2 - T_1) \qquad \text{(EQ. 3)}$$

The true three dimensional solution of heat flow is complex, requiring the use of partial differential equations and vector analysis. Linear heat flow is assumed in most cases to achieve a first order solution to a thermal problem.

Electrical Power, P in watts (joules per second) causes the heating in an electrical component. Power input is normally defined as voltage times current and instantaneous power, p(t) is instantaneous voltage times instantaneous current.

Electrical Energy, E in units of joules or watts·seconds is the time (t) integral of power,

$$E = \int_0^t p(t)\,dt \qquad \text{(EQ. 4)}$$

Since Heat Flow, Q is the energy rate in joules per second which, by definition, is watts and input power is also in watts, the power supplied to an active element must reach equilibrium with the rate of heat flow being conducted or radiated away. As such, Q (power) is used as a symbol for heat flow away from a junction or device to separately distinguish this energy rate from the applied electrical Power, P at the junction. It is important to note that the temperature, ΔT must rise as a driving force to establish this equilibrium. Until equilibrium is established, the heat capacity of the body determines the transient period of change.

When heat is generated by the output driver, it changes the parameters of the device model. As an example, for the bipolar cds Spice hybrid-π model shown in Figure 3, an added temperature node illustrates the interactive temperature effect on the parameters of the device model. [5] Power supplied to the transistor generates heat flow in the form of an equivalent current generator, Q. The g_{tb} and g_{tc} current sources are driven by the ΔT heat source. While similar models are not available for the Power BiMOS output devices, a device model for temperature is being developed.

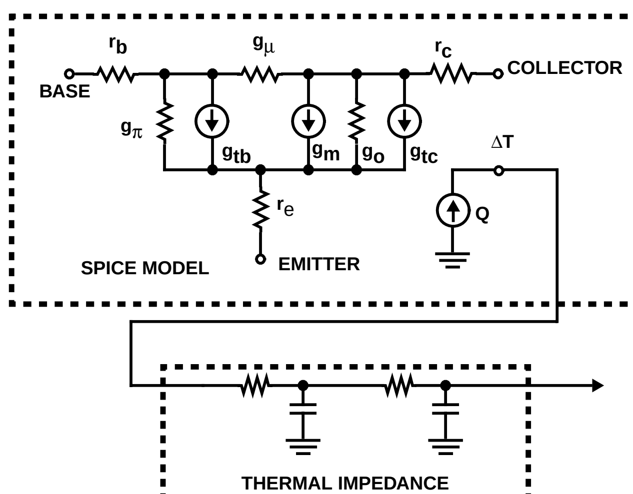

FIGURE 3. MODIFIED HYBRID-π SPICE MODEL TO SHOW THE POWER TO TEMPERATURE DRIVE SOURCE

Thermal Conductivity and Thermal Resistance

Fourier's differential equation defines the temperature gradient for an incremental quantity of heat, dT flowing through a body in a linear unidirectional path of incremental length dx and cross-sectional area, A as:

$$Q = -kA\left(\frac{dT}{dx}\right) \quad \text{(EQ. 5)}$$

Separating variables and integrating temperature, T from T_1 to T_2 over the heat flow path, d from x_1 to x_2, gives the linear solution,

$$Q = \frac{kA}{d}(T_1 - T_2) \quad \text{(EQ. 5A)}$$

Thermal conductivity, k is a well documented material parameter with a definition of heat flow conducted per ^oC normal to the direction of heat flow through a unit distance per unit area of material. While the coefficient of thermal conductivity is a physical parameter correctly expressed in SI units as W/m • ^oC or (W/^oC) • (m/m^2), it is more commonly listed in reference publications as W/cm • ^oC or Btu/fthr • ^oF.

Thermal resistance, R_θ is more frequently used to describe the capability of heat transfer through a given body of fixed dimensions. The units of thermal resistance are ^oC/W. For a composite body of material, the thermal resistance equation becomes:

$$R_\theta = \frac{d}{kA} = \frac{\Delta T}{Q} \quad \text{(EQ. 6)}$$

which defines thermal resistance, R_θ in relation to temperature difference and heat flow. Since ΔT represents the driving force or potential for heat flow and thermal resistance determines the rate of heat flow conduction, we may consider the equation:

$$\Delta T = Q R_\theta \quad \text{(EQ. 6A)}$$

as analogous to the Ohms Law equation, V = IR. This analogy is similar to a node equation voltage difference with temperature difference as the driving force.

Since we cannot directly adjust temperature in an electrical circuit, temperature and heat flow are affected only by the thermal resistance and the level of power input. As such, we may directly substitute electrical power dissipation, P_D for its heat equivalent, Q to establish the equation,

$$\Delta T = P_D \times R_\theta \quad \text{(EQ. 6B)}$$

Where T_J is junction temperature and T_C is case temperature, $\Delta T = (T_J - T_C)$. The symbol, R_θ, is used here to represent the thermal analogy to electrical resistance. In the material that follows, thermal resistance is defined by the standard symbol now in use, which is θ_{JC} or θ_{JA}, to represent the junction-to-case or junction-to-ambient paths. Therefore,

$$T_J - T_C = P_D \times \theta_{JC} \quad \text{(EQ. 6C)}$$

Thermal resistance of a well defined structure may be determined, given conditions of unidirectional and uniform heat flow using equation Equation 6. The thermal conductivity coefficients in Table 2 are intended as guidance to understand and improve the thermal resistance of IC heat sink structures and can be used for an approximation for the θ_{JC} of the IC package. However, the complex physical configuration of most ICs require direct measurement as the only practical way to determine thermal resistance. Even where direct measurement can provide θ_{JC} or θ_{JA} values for a given IC, the final application θ_{JA} is still quite dependent on the operating environment. The IC and heat sink structure should always be given a final equipment design check measurement over the operating temperature range.

Figure 2 illustrates the Junction-to-Ambient thermal paths for the 15 Lead Single-in-Line (SIP) Power IC package. The Junction-to-Case thermal resistance, θ_{JC}, for this package is primarily the sum of the thermal resistance values for the die, die attach and tab of the IC. For an assumption of linear heat flow, each component path of the IC has a defined thermal resistance. Heat generated at the surface of the die flows through the silicon die to the copper heat sink tab and external heat sink. Heat also flows through the lead connections and through the molding of the plastic package to the surface of the IC. The combined thermal resistance paths through the plastic and leads is significantly greater than the direct die-to-tab path in the IC.

Other methods used for maintaining junction temperature control employ the heat conducting qualities of the copper lead frame. The Power PLCC package of the HIP0080, conducts heat through the lead frame structure to the IC ground leads and then to the copper ground area on the external PC Board as illustrated in Figure 4. The θ_{JC} for this package actually refers to the Junction-to-Leads path with 14^oC/W thermal resistance which is quite low compared to a standard PLCC package. For this package, the PC Board heat sink area is the most significant part of the Junction-to-Ambient, θ_{JA} thermal resistance.

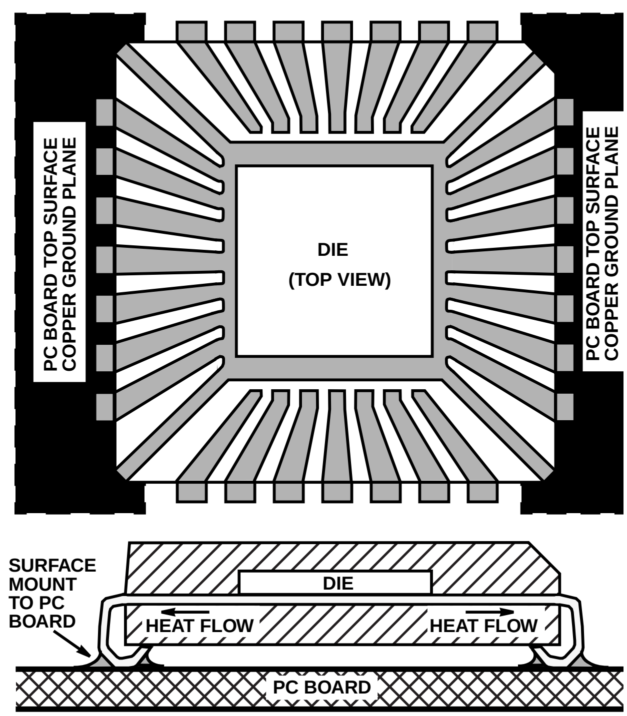

FIGURE 4. THE HIP0080 PLCC PACKAGE ILLUSTRATING THE COPPER LEAD FRAME CONSTRUCTION DESIGNED TO CONDUCT HEAT TO THE GROUND LEADS

The actual field pattern of heat flow lines is similar to that of an electric field having curving divergent lines. As heat diffuses from its source of excitation, the isothermal lines of the gradient are orthogonal to the direction of heat flow. To provide an example that illustrates how thermal resistance is determined, we can assume a simpler linear heat flow model. The direct path model of Figure 5A should give a reasonable approximation for θ_{JC} of the IC in Figure 2 by assuming linear heat diffusion in the shape of a truncated pyramid, spreading at 45^o away from the active junction area of the die. Then, for an area approximation, the average length and width dimension of each material in the pyramid is increased by an amount equal to thickness.

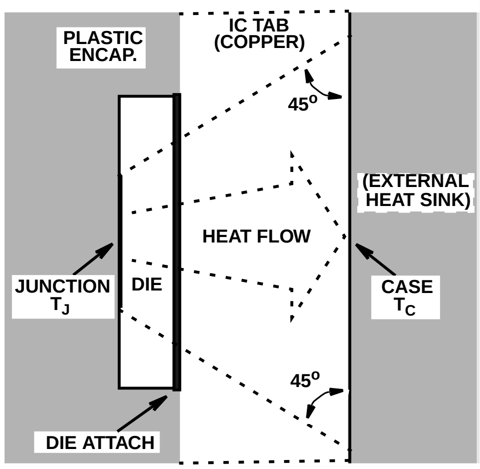

FIGURE 5A. AN IC CROSS-SECTION USING A 45° DIFFUSION PATTERN (IN THE SHAPE OF A TRUNCATED PYRAMID) TO ILLUSTRATE BY APPROXIMATION, THE HEAT FLOW FROM THE JUNCTION TO THE CASE FOR THE IC PACKAGE SHOWN IN FIGURE 2

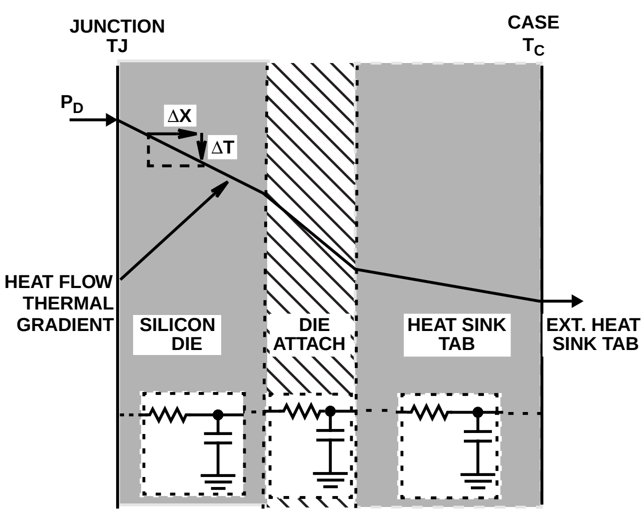

FIGURE 5B. HEAT FLOW FROM JUNCTION-TO-CASE SHOWING THE THERMAL GRADIENT, ΔX/ΔT. RESISTOR AND CAPACITOR SYMBOLS REPRESENT THERMAL RESISTANCE AND HEAT CAPACITY AS COMPONENTS IN THE HEAT FLOW PATH

FIGURE 5.

Example: Using Equation 6 for thermal resistance and the suggested model in Figure 5A, we can make the following calculations for thermal resistance.

1. For active dissipation area on the die, assume a junction area of 54 x 89 square mils and, for a die thickness of 20 mils, the heat diffuses out at 45^o to increase the effective dimension in each direction by 2 x 1/2 the die thickness or 20 mils. As such, the effective die dimensions are:
   A = 74 x 109 = 8066 square mils or 5.204 x 10^{-6} square meters
   d = 20 mils = 50.8 x 10^{-5} meters
   Using the value for thermal conductivity for silicon from Table 2, and applying to Equation 6, we have:
   $$R_\theta(\mathrm{Die}) = \frac{d}{kA} = \left(\frac{5.08 \times 10^{-4}}{84 \times 5.2 \times 10^{-6}}\right) = 1.16^\circ\mathrm{C/W}$$
2. For the 1 mil thick Die Attach Solder, the dimensions are approximately equal to the that of the junction plus 2 times the die thickness plus 2 x 1/2 the Die Attach thickness or 40 + 1 = 41 added for each dimension.
   Then:
   A = 95 x 130 = 12350 square mils or 7.968 x 10^{-6} square meters
   d = 1 mil = 2.54 x 10^{-5} meters
   For Lead, k = 35W/m•^{o}C
   R_θ (Die Attach) = 0.09^o C/W
3. For the 60 mil Copper Heat Sink Tab that is an integral part of the IC, the Tab dimensions equal the active junction area dimensions plus the 45^{o}spread in each direction or (2 x 20) + (2 x 1) + (2 x (1/2) x 60) = 102 mils
   A = (54 + 102) x (89 + 102) or 156 x 191 mils,
   A = 2.980 x 10^{+4} square mils = 19.22 x 10^{-6} square meters
   d = 60 mils = 1.52 x 10^{-3} meters
   k = 381 W/m•^{o}C
   R_θ (Tab) = 0.21^{o}C/W.

The calculated thermal resistance is the sum of the series thermal resistance values which are:

θ_{JC} = 1.16 + 0.09 + 0.21 = 1.46^{o}C/W.

Based on measurement, the actual junction-to-case thermal resistance rating for to this package is 3^{o}C/W. The active junction area used is the approximate dimension for one output driver of the HIP0082. With added active drivers on the chip there will not be a proportional decrease in the thermal resistance due to an overlap in the assumed 45^o diffused fan-out for heat flow.

The temperature gradient changes as heat flows from the junction to the case of the IC. The slope of the gradient changes with changes in the thermal resistance of each material. This effect is illustrated in Figure 5B.

Heat Capacity and Transient Thermal Impedance

Heat Capacity, C_T is the amount of heat required to raise the temperature of a mass (body or component part) by 1^{o}C. When the mass of the body is known, we may determine the specific heat (abbreviated sp. ht.). As such, the heat capacity of a body is calculated to provide tabulated values for specific heat using Equations 1 and 3, where:

$$c = \frac{C_T}{m_k} = \frac{Q}{m_k(\Delta T)}$$

When gram-calorie units of specific heat, c are given in cal/gram (°C), the conversion to Joules, J is:

$$\text{sp. ht., c in } \left(\frac{\text{cal}}{\text{gram}(^\circ C)}\right) \text{Units} = 4.186\text{J}$$

Also, 1 kilogram-calories = 4.186kJ

Then, $C_T = c \times m_k = 4.186 \times (\text{sp. ht.}) \times \text{wt}$.
where the weight, wt = Density, D x Volume, V.
(See Table 2 for sp. ht. values of various materials).

In Figure 5B, an electrical analogy of series resistance and shunt capacitance is used to illustrate the electrical equivalence of thermal resistance and heat capacity. A short duration power pulse at the chip junction integrates the power as heat energy which will diffuse to eventually be detected on the case (tab) of the IC. Since the RC filter circuit is an integrating network, this is a reasonable symbol analogy using the comparison of electrical current to heat flow units of joules/second or watts.

Calculation of the heat capacity and the thermal time constant can be made from assumptions similar to those used for the unidirectional analysis of thermal resistance. However, since the thermal capacitance is distributed in the material, a computer simulation for a ladder of small distributed RC elements should be used for an accurate result. As a first order approximation, we can assume strict unidirectional heat flow (without side diffusion) to determine the thermal time constant of the silicon die.

Example: Using centimeters (cm) and grams (g) for units of length and mass, calculate the thermal time constant given a fixed volume of silicon with Area, A = x•y and die thickness, d = 20 mils = 0.0508cm. The thermal resistance is:

$$R_\theta = (1/k)\ \text{cm}^oC/W \bullet (d/x \bullet y)\ \text{cm/cm}^2.$$

Using the formula derived value for k given Note 1, Table 2; at 120^{o}C the conductivity, k = 0.98 W/cm^{o}C

$$R_\theta = [(1/.98)\ \text{cm}^oC/W] \bullet [(d/x \bullet y)\ \text{cm/cm}^2]$$
$$= (1.02 \bullet d/x \bullet y)\ ^oC/W.$$

For Heat Capacity,

$$C_T = 4.186 \bullet c \bullet wt = 4.186 \bullet c \bullet D \bullet V.$$

where V = (d•x•y) cm^3 and silicon density, D = 2.41 g/cm^3.

Then,

$$C_T = [(4.186 \bullet 0.181)\ J/g^oC] \bullet [(2.41)\ g/cm^3] \bullet [(d \bullet x \bullet y)\ cm^3]$$
$$= (1.826 \bullet d \bullet x \bullet y)\ J/^oC.$$

Combining C_T and R_θ for the thermal time constant,

$$R_\theta C_T = [(1.02 \bullet d/x \bullet y)\ ^oC/W] \bullet [(1.826 \bullet d \bullet x \bullet y)\ J/^oC]$$
$$= 1.862 \bullet d^2\ \text{ms}.$$
$$= 1.862 \bullet (0.0508)^2 = 4.81\text{ms}.$$

The actual value will be less due to the distributed capacitance effect. While the physical shape of each component

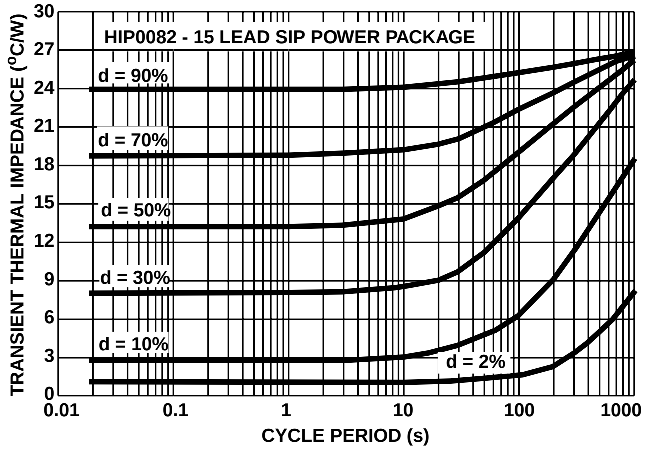

FIGURE 6A. TYPICAL CHARACTERISTIC PLOT OF THE TRANSIENT THERMAL IMPEDANCE VERSUS CYCLE PERIOD (1/f) FOR THE HIP0082 QUAD LOW SIDE POWER DRIVER

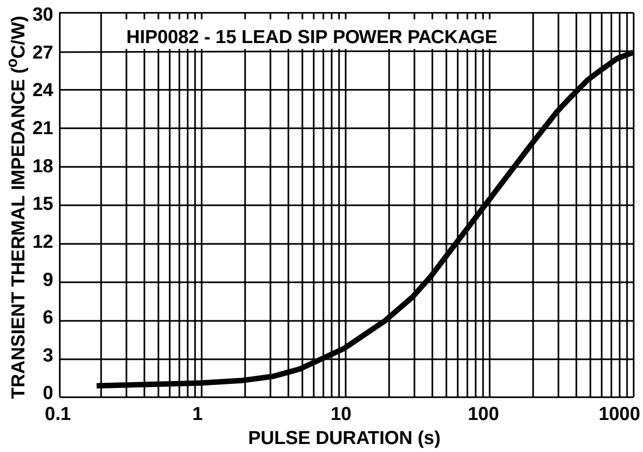

FIGURE 6B. TYPICAL CHARACTERISTIC PLOT OF THE TRANSIENT THERMAL IMPEDANCE FOR A SINGLE PULSE VERSUS PULSE DURATION FOR THE HIP0082 QUAD LOW SIDE POWER DRIVER

FIGURE 6.

section in an IC is generally too complex for simple calculations, evaluations can made using computer aided design, including methods used for finite element analysis and electrical circuit design, including methods used for fiite element analysis and electrical circuit design.

Direct measurement can give us a composite transient thermal impedance curve and the thermal time constants. Power is normally applied to the IC junction as pulsed energy with variations of duty cycle and frequency. From the test data, we can derive an equivalent RC time constant by plotting the effects of transient energy change into the IC. As an example, test results for an application of the HIP0082 are illustrated in Figures 6A and 6B. The data was taken with the HIP0082 mounted on a copper PC Board.

In Figure 6A, Transient Thermal Impedance versus Cycle Period is plotted for a family of 6 curves with a duty cycle variation from 2%, to 90%. For low duty cycles and small cycle periods, the Transient Thermal Impedance, Z_θ is very low. For very long cycle periods and large duty cycles, Z_θ

approaches the static thermal resistance value of θ_{JA} that includes the IC and PC Board, approximately $27^oC/W$. A similar curve is plotted in Figure 6B to show the transient thermal impedance versus pulse duration for a single non-repetitive pulse. The thermal time constants are empirically derived from the log-linear slope of the curves.

For low duty cycle and single pulse conditions the thermal impedance is quite low at ~1Ω. This does not imply that we can have the best of both worlds by supplying all needs with short power pulses. It is important to realize that short duration pulses can inject high instantaneous peak power while the measurement of thermal impedance remains small. Localized heating does occur at the junction in proportion to the applied power level and can cause stress damage at the junction before it is detected by external monitors.

The transient thermal impedance time constants shown in Figures 6A and 6B are much longer than the $R_\theta C_T$ of 2.33 ms previously calculated. The time constants represented by the change of curve slope in the range of 1s to 1000s corresponds to the IC copper tab and the external PC board. The transient thermal impedance curve of Figure 6B indicates that a single pulse width up to several seconds will not give a significant indication of heat at the chip junction. The thermal time constant of the package delays heat flow to case. Until the junction temperature rises sufficiently to force a sustained heat flow, the “apparent” thermal impedance of the package will continue to appear low.

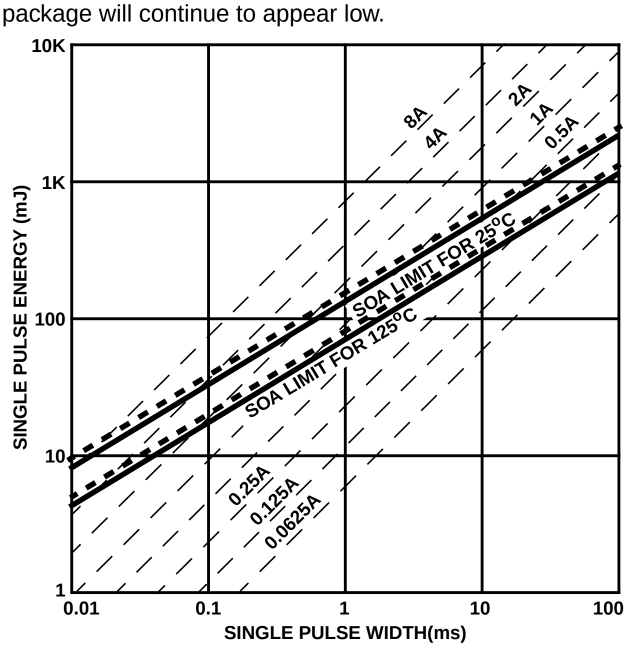

FIGURE 7. RATING LIMITS FOR THE HIP0082 OUTPUT CLAMP FOR ENERGY vs THE WIDTH VARIATION OF A SINGLE CURRENT PULSE INTO THE CLAMP. THE SOA LIMIT IS THE MAXIMUM SAFE LEVEL FOR SINGLE PULSE ENERGY AND SHOULD NOT BE EXCEEDED

Power switching output circuits with internal clamp protection require rating protection for short duration peak power pulses. As such, conditional limits are placed on the maximum energy of a single pulse. The rating curve of Figure 7 illustrates the Energy vs Time single pulse capability of the HIP0082 which has an internal drain-to-gate clamp protection circuit. The Safe Operating Area (SOA) limit was determined by applying pulse width variation and increasing the pulse current to a point of stress. The data was taken in an IC socket without a heat sink attached. The SOA limit is noted to be lower for high ambient temperatures. Peak currents are also noted for referenceThe product of clamp voltage (90V specification maximum) times the pulse current is peak dissipation in Watts. The product of watts and milliseconds (ms) is energy in millijoules (mJ). The single pulse Energy equation is:

$$\mathrm{Energy} \approx 90V \times \mathrm{Current} \times \mathrm{Time}$$

The Figure 7 rating curve is based on well defined energy pulses. However, it does have good correlation to the flyback energy pulse generated when an inductive load is switched off. The collapsing energy of an inductor forces an exponential decay in the current waveform. The integrated energy of the pulse times the period of field collapse provide the basis for correlation and are discussed a later section.

Maximum Temperature Ratings

A general equation for dissipation should express the total power dissipation in a package as the sum of all significant elements of dissipation on the chip. However, in Power BiMOS Circuits very little dissipation is needed to control the logic and pre-driver circuits on the chip. The overall chip dissipation is primarily the sum of the I^2R dissipation losses in each NMOS output channel, where I is the output current and R = is the channel resistance, R_{DSON}. The dissipation, P_D in each output driver is:

$$P_D = I^2 \times R_{DSON} \qquad \text{(EQ. 7)}$$

The total power dissipated for k = 1 to n output drivers:

$$P_D = \sum_{k=1}^{n} P_k \qquad \text{(EQ. 8)}$$

Which, as an expression, sums the dissipation all output drivers without regard to uniformity of dissipation in each MOS channel.

The current, I is the determined by the output load when the channel is turned ON. The channel resistance, R_{DSON} is determined by the IC design, the level of forward gate voltage drive at Turn-ON and the on-chip temperature at the conducting output. While the effective value of R_{DSON} may include resistance in the interconnecting metal on the chip and the bond wires of the package, this is assume to be small for the range of currents discussed here.

Other sources of dissipation in the IC are the result of switched inductive loads. The energy stored in the Inductor is generally defined in millijoules (Watt-seconds) and, for a given current, is equal to:

$$E_L = \frac{LI^2}{2} \qquad \text{(EQ. 9)}$$

which is dissipated in the output driver when the inductive flyback pulse is clamped internally in the IC. This is discussed in a later section.

As noted in Equation 6C, the temperature rise in the package due to the power dissipation is the product of the dissipation, P_D and the θ_{JC} of the package. For a junction temperature, T_J , and a case temperature, TC;

$$T_J = T_C + P_D(\theta_{JC}) \quad \text{(EQ. 10)}$$

or, given that T_C should not exceed a known temperature, we may express the equation as:

$$T_C = T_J - P_D(\theta_{JC}) \quad \text{(EQ. 10A)}$$

Since this solution relates only to the package, further consideration must be given to a practical heat sink. The equation of linear heat flow assumes that the Junction-to-Ambient thermal resistance, θ_{JA} is the sum of the Junction-to-Case thermal resistance, θ_{JC} and the Case-to-Ambient thermal resistance θ_{CA}. For a power device such as illustrated in Figure 2, the Case-to-Ambient thermal resistance is the external Heat Sink. The Junction-to-Ambient thermal resistance, θ_{JA} is the sum of all thermal paths from the chip junction to the environment with an ambient temperature, T_A and can be expressed as:

$$\theta_{JA} = \theta_{JC} + \theta_{CA} \quad \text{(EQ. 11)}$$

The Junction-to-Ambient equations equivalent to Equations 10 and 10A are:

$$T_J = T_A + P_D(\theta_{JA}) \quad \text{(EQ. 12)}$$

and,

$$T_A = T_J - P_D(\theta_{JA}) \quad \text{(EQ.12A)}$$

Where the ambient temperature and dissipation conditions are subject to user control, we can use Equation 12 and 12A to determine the junction temperature rise or maximum ambient temperature limits.

When the maximum junction and ambient temperatures are known, Equation 12A gives the solution for maximum allowed IC dissipation, P_D as:

$$P_D = \frac{(T_J - T_A)}{\theta_{JA}} \quad \text{(EQ. 13)}$$

Example 1: For the HIP0082 package, θ_{JC} = 3^{o}C/W and the worst case junction temperature, as an application design solution, should not exceed 150^{o}C. For a given application, Equation's. 7 and 8 determine the dissipation, P_D. Assume the package is mounted to a heat sink having a thermal resistance of 6^{o}C/W. And, for the given application, assume the dissipation is 3W and the ambient temperature (T_A) is 100^{o}C. From Equation 11, θ_{JA} is 9^{o}C/W. The solution for junction temperature using Equation 12 is:

$$T_J = 100°C + [3W(9°C/W)] = 127°C$$

Example 2: Assume for the HIP0080, θ_{JA} = 30^{o}C/W when mounted on a PC Board with good heat sinking characteristics. Again, the worst case junction temperature, as an application design solution, should not exceed 150^{o}C. Assume for this application, the dissipation, P_D = 1.5W. Given a maximum junction temperature of 150^{o}C, we can determine the maximum allowable ambient temperature from Equation 12A as follows:

$$T_A = 150°C - [1.5W(30°C/W)] = 105°C$$

Example 3: For the HIP0080 mounted on a PC Board with a θ_{JA} = 35^{o}C/W, determine the maximum dissipation for an ambient temperature of T_A = 125^{o}C. Using Equation 13:

$$P_D = \frac{(T_J - T_A)}{\theta_{JA}} = \frac{(150 - 125)}{35} = 0.714W$$

Dissipation Derating Curves

The general solution of Equations 10 through 13 can be easily adapted to a graphical illustration as shown in the commonly used "Derating Curve" of Figure 8. The θ_{JC} rating of the HIP0082 Power Package is 3^{o}C/W with a substantially higher value for θ_{JA}. The junction-to-case thermal resistance, θ_{JC} is based on temperature measurements at the case to external heat sink interface. Practical considerations require additional case-to-air, θ_{CA} heat sinking to achieve an overall low θ_{JA}. This is shown in the middle curve of the Figure 8 graph where the 3^{o}C/W heat sink for θ_{JC} and 6^{o}C/W for the external heat sink gives an overall θ_{JA} of 9^{o}C/W.

It should also be noted that the "derating factor" is the reciprocal of the device thermal resistance. Dissipation Derating is commonly defined in milliwatts per ^{o}C and is done to protect the maximum junction temperature in high ambient temperature conditions. The allowed dissipation is reduced to zero when T_J = 150^{o}C. Therefore the derating curve intercept point for P_D = 0 is at the maximum junction temperature of 150^{o}C.

The junction-to-ambient thermal resistance, θ_{JA} is more correctly defined for the conditions of the application environment. Normally, a reduction in the thermal resistance can be achieved by increasing the square area size of the external heat sink such as to conduct more heat away from the package. Where the heat conducting path from the junction is through the lead frame, to the package pins, added copper ground area on PC Board is needed to reduce θ_{JA}.

Current Limited, Maximum Dissipation Curves

Achieving maximum performance from Power BiMOS Drivers with Multiple Outputs requires careful attention to the maximum current and dissipation ratings. While each output has a maximum current specification consistent with the device structure, all such devices on the chip may not be simultaneously rated to the same high current level. In such cases, the total current and dissipation on the chip must be within the maximum allowable ratings. For many applications, all output drivers on a chip are equally loaded. Given equal loading, the following discussion is intended to provide a relatively simple method to calculate the boundary conditions for maximum allowed dissipation and current.

For equal current loading in each output driver, assuming saturated Turn-ON; the maximum current versus ambient temperature may be graphically plotted. To derive the curve parameters for given ratings, we need to substitute the expression for dissipation from Equation's. 7 and 8 into the junction temperature Equation 10A. Where the currents from each output driver are equal (i.e. $I = I_1 = I_2 ..... = I_k ..... = I_n$), we have the following maximum current solution based on total dissipation:

$$P_D = n(P_k) = n\,(I^2 R_{DSON}) = \frac{(T_J - T_C)}{(\theta_{JC})} \quad \text{(EQ. 14)}$$

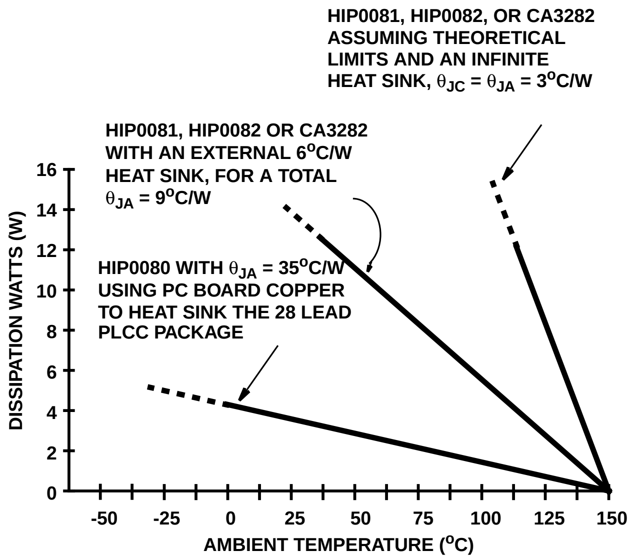

FIGURE 8. DISSIPATION DERATING CURVES SHOWING MAXIMUM DISSIPATION vs AMBIENT TEMPERATURE. THE SLOPE OF EACH CURVE IS THE RECIPROCAL OF THERMAL RESISTANCE FOR THE CONDITIONS NOTED AND THE SAFE OPERATING AREA FOR EACH CURVE IS BELOW THE LINE

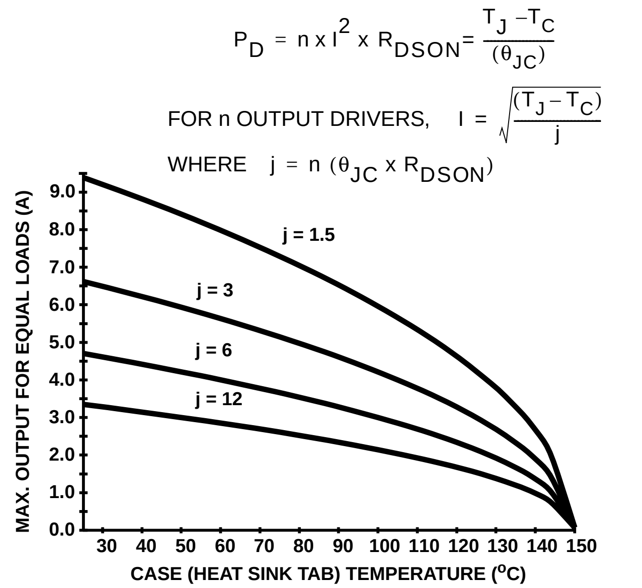

FIGURE 9. GENERAL GRAPH SOLUTION FOR MAXIMUM OUTPUT DRIVER CURRENT, I vs CASE TEMPERATURE, T_C GIVEN EQUAL LOAD CURRENTS AND A MAXIMUM JUNCTION TEMPERATURE, $T_J = 150^oC$

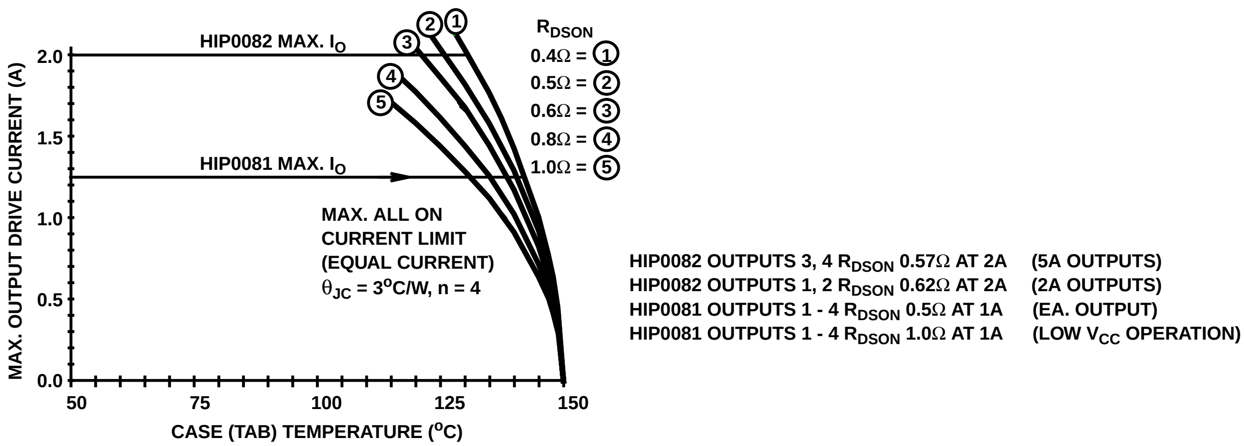

FIGURE 10. HIP0081 AND HIP0082 MAXIMUM CURRENT vs CASE TEMPERATURE. ALL 4 OUTPUTS ON WITH EQUAL CURRENT

Solving for current, I gives:

$$I = \sqrt{\frac{(T_J - T_C)}{n\theta_{JC}R_{DSON}}} \qquad \text{(EQ. 15)}$$

The number of output drivers ON and conducting may be from n = 1 to 4 for quad drivers such as the HIP0081 and HIP0082 or n = 1 to 8 for an octal driver such as the CA3282. Maximum temperature, dissipation and current ratings must be observed. For n equal conducting output drivers, we can now plot the maximum rated current, I vs the maximum case temperature, T_C.

The denominator term in the radical of Equation 15, $n\theta_{JC}R_{DSON}$, is used to plot a set of normalized curves, as shown in Figure 9. Using the HIP0081 and HIP0082 as examples, Figure 10 illustrates the boundaries for temperature and current for a range of R_{DSON} with all four outputs ON and equal output current. The curve coordinates are output current, I vs Case Temperature, T_C. The maximum total current limits of 5A for the HIP0081 and 8A for the HIP0082 directly relate to the Absolute Maximum Ratings as defined in the datasheet.

Inductive Load Switching - Pulse Energy And Dissipation Calculations

To expand on the subject of Inductive Loads, it is necessary to find the solution for energy loss that results from active switching of inductive load current, including the dissipation from the flyback pulse when the output is turned off. When the energy loss is found, duty factor and frequency variables may be applied to calculate dissipation.

The time integral of instantaneous power is used to determine energy during Turn-ON and Turn-OFF. The indicated solution is given in Equations 16 and 16A where the instantaneous power, $P(t) = V(t)I(t)$ or $I(t)^2R$, is integrated over a given time period. The answer is defined in joules (watt-seconds) and the resistance, R is the series resistance in which the energy, E is to be calculated.

$$E\Big|_0^t = \int_0^t V(t)\, I(t)dt \qquad \text{(EQ. 16)}$$

$$\text{or } E\Big|_0^t = \int_0^t I(t)^2\, R dt \qquad \text{(EQ. 16A)}$$

The general solution of Turn-ON and Turn-OFF for an RL circuit is a well documented transient differential equation. The expression for voltage is usually defined for a fixed step function but may be more complex.

For the illustrated conditions shown in Figure 11, we have a low side MOS driver with an RL load. The initial condition for current at time t = 0 may be other than zero. To assure that an inductive kick pulse (flyback) at Turn-OFF does not exceed the ratings of the MOS output driver, a zener diode feedback circuit is used to clamp the output pulse. The zener diode is used as a feedback clamp to turn on the output stage when the zener diode threshold voltage, V_Z plus the gate threshold voltage, $V_{GS(TH)}$ is exceeded.

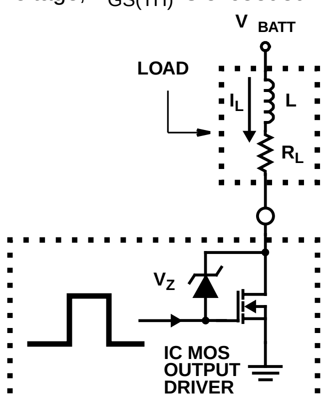

FIGURE 11A.

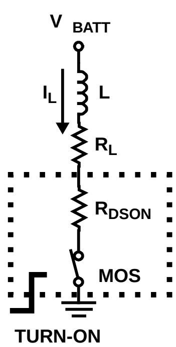

FIGURE 11B. EQUIVALENT ON STATE CIRCUIT

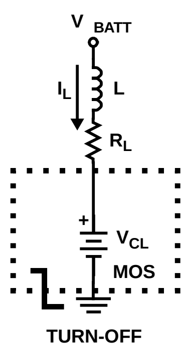

FIGURE 11C. EQUIVALENT OFF STATE CIRCUIT DURING INDUCTOR DISCHARGE

FIGURE 11. TYPICAL MOS DRIVER OUTPUT CIRCUIT WITH A SWITCHED INDUCTIVE LOAD. AN INTERNAL ZENER DIODE VOLTAGE CLAMP IS USED FOR FLYBACK OVER-VOLTAGE PROTECTION

The circuit shown in Figure 11 is a common method used to limit the flyback voltage and is used in the CA3282, HIP0080, HIP0081 and HIP0082 Low Side Drivers. Since the power dissipated in a pulse clamped mode of operation is dissipated in the IC, a determination of the pulse energy is needed to protect the rating limits of the IC. The following solutions used here are general and may be adapted to any of the Power BiMOS IC types.

The general equation for instantaneous transient current in the series loop of a step switched inductive load is:

$$I_L(t) = I_O(1 - Ae^{-t/T}) \qquad \text{(EQ. 17)}$$

where $\quad A = \left(1 - \dfrac{I_B}{I_O}\right)$

Given the proper values for the initial conditions, this solution applies to either Turn-ON or Turn-OFF of an inductive load. The Initial Current at t = 0 is I_B and the Steady State Current, for $t \gg T$, is I_O. The circuit time constant, T is L/R where R is the total series resistance in the loop and L is the inductance. Since different Turn-On and Turn-OFF equation parameters apply, the initial conditions differ for each case. For example, the loop value of R is different in the Turn-ON and Turn-OFF periods. While there may or may not be an initial current at Turn-ON, the Turn-OFF condition has an implied initial current.

Inductive Load Turn-On

When the MOS input is switched ON, the I^2R expression for instantaneous power may be used for the integral solution of energy loss in R_{DSON} of the Output Driver. At Turn-ON, the general solution of energy loss in the series resistance, R; is found by substituting Equation 17 into Equation 16A. The integral equation is:

$$E_{RON} = \int_0^t I_O^2\left(1 - Ae^{-\frac{t}{T}}\right)^2 R dt \qquad \text{(EQ. 18)}$$

When the squared term is expanded and integrated, the solution for Turn-ON Energy in the resistance, R becomes

$$E_{RON} = I_O^2 R\left(t + 2ATe^{-\frac{t}{T}} - \frac{A^2T}{2}e^{-\frac{2t}{T}}\right)\Bigg|_0^t \qquad \text{(EQ. 18A)}$$

The current, $I_L(t)$ flows in the MOS output and, over time, dissipates power in the drain-to-source channel resistance, R_{DSON}. Since the Equation 18A is a general solution of energy loss in the series resistance of an RL circuit, we may treat R as the sum of all series resistance in the RL loop. And, if needed to limit current in the load, additional resistance may be added in series external to the IC and Coil as part of the resistive load.

If $R = R_L + R_{DSON}$, the E_{RDSON} energy loss in an IC output is the R_{DSON} term expression,

$$E_{RDSON} = I_O^2 R_{DSON}\left(t + 2ATe^{-\frac{t}{T}} - \frac{A^2T}{2}e^{-\frac{2t}{T}}\right)\Bigg|_0^t \qquad \text{(EQ. 19)}$$

We may use Equation 19 to find the complete Turn-ON energy, E_{RDSON} loss in each MOS output driver.

Note that the time limit is not restricted and may continue to apply for the steady state ON condition. Given the following symbol definitions and initial conditions, the complete energy solution may now be calculated. For Figure 11, the Turn-ON circuit parameters are:

$I_L(t)$ Instantaneous Inductor Load Current
V_{BATT} Output Load Power Supply
V_{CL} Pulse Clamp Voltage, $V_{CL} = V_Z + V_{GS(TH)}$
L, R_L Load Inductance and Resistance
T Time Constant, $T = L/R$ where $R = R_L + R_{DSON}$
I_O Steady State Current, $I_O = \frac{V_{BATT}}{R}$ for $t \gg T$
I_B Initial Current flowing at t = 0.
R_{DSON} MOS ON State Channel Resistance
P_{DON} Steady State Dissipation, $I_O^2 R_{DSON}$

Expanding Equation 19, the complete expression for Turn-ON energy loss in the IC over time t = 0 to t is:

$$E_{RDSON} = P_{DON}\left[\left(t + 2ATe^{\frac{-t}{T}} - \frac{A^2T}{2}e^{\frac{-2t}{T}}\right) - \left(2AT - \frac{A^2T}{2}\right)\right]$$

$$= P_{DON}\left(t - 2AT\left(1 - e^{\frac{-t}{T}}\right) + \frac{A^2T}{2}\left(1 - e^{\frac{-2t}{T}}\right)\right) \quad \text{(EQ. 20)}$$

An approximation solution for E_{RDSON} is impractical because the energy value will continue to increase with time. However, as the time is extended, we may assume that $e^{-t/T}$ terms of Equation 20 will diminish to a small value and further reduce the complexity of the Turn-ON calculation. By substituting t = kT, where k gives us a scaled value of T, and then increasing k to a large value reduces the exponential terms to zero. If the initial current, I_B is zero, then A = 1. After making these substitutions in Equation 20, we have:

$$E_{RDSON} = P_{DON}T\left[k - 2 + \frac{1}{2}\right] = (k - 1.5)\,T\,P_{DON} \quad \text{(EQ. 20A)}$$

which applies for $t \geq 5T$, i.e. $k \geq 5$.

The Equation 20A result indicates that we could have calculated the steady state energy for a kT period of time and then subtracted $1.5(TP_{DON})$ to account for the reduced energy during the transient Turn-ON interval. We may use Equation 20A for a short cut calculation, but only for time periods greater than 5T. The error is less than 1% for $t \geq 4.7T$.

Example: Given the following data and using Equations 17 and 20, calculate $I_L(t)$ and E_{RDSON}.

$V_{BATT} = 13.5V$

$V_{CL} = 82V$

$L = 100mH$

$R_L = 9.5\Omega$

$R_{DSON} = 0.5\Omega$

$R = R_L + R_{DSON} = 9.5 + 0.5 = 10\Omega$

$I_O = V_{BATT}/R = 13.5/10 = 1.35A$

$I_B = 0$

$A = 1$

$T = L/R = 100/10 = 10ms$

Let t = 5T = 50ms

Then, $I_L(t) = 1.341A$ at t = 5T

And $E_{RDSON} = 32.02$ mJ.

For the Short Cut method, using Equation 20A
$P_{DON} = R_{DSON}I_O^2 = 0.5 \times (1.35)^2 = 0.911W$

For k = 5;

$E_{RDSON} = (k-1.5)T\,P_{DON} = 3.5TP_{DON} = 31.89mJ$.

For switching conditions such as may be encountered when driving injectors and motors, total energy increases with frequency. Also, for large inductors with continuous switching, Turn-ON may not be completed before Turn-OFF; making the I_B term significant and the Turn-ON Energy much larger.

Inductive Load Turn-off

The Turn-OFF Energy, E_{CLAMP} in the MOS output may add a significant contribution to the power dissipation in the IC. Also, short duration pulse energy must be determined for a safe rating check. Turn-OFF current may be calculated using Equation 17, given a value for the instantaneous current, I_B as an initial condition at switch-off. In the clamp mode, the MOS drain voltage is fixed at V_{CL} and may be regarded as a fixed voltage drive source, but only while the clamp is sustained. The degree of MOS turn-on varies according to the sustaining needs of the clamp. For this reason, the I^2R approach used for Turn-ON does not apply. The best approach for the Turn-OFF energy solution is to use the P(t) = V(t)I(t) version of the integral power equation where $V(t) = V_{CL}$ during the clamp period. The Turn-OFF energy equation is:

$$E_{CLAMP} = \int_0^t (V_{CL} I_L(t))\,dt \quad \text{(EQ. 21)}$$

Since current must be continuous during switching, the current flowing in the inductor at switch-off becomes the Initial Current, I_B in the current equation. To compensate for the switching discontinuity, a flyback voltage pulse proportional to the differential rate of current change is generated. For an unclamped pulse,

$$V(t) = -L\left(\frac{di}{dt}\right)$$

The flyback pulse may be quite large, requiring a clamp circuit for over-voltage protection.

For the equivalent clamp circuit shown in Figure 11, Turn-OFF occurs in 3 stages:

1. Following turn-off, the flyback pulse quickly increases to the clamp voltage level. A small incremental loss occurs before the flyback pulse is clamped. However, this is assumed to be negligible because the Power BiMOS Output drivers typically switch off in less than 10µs.
2. The flyback voltage is clamped by the output clamp circuit until the energy stored in the inductor is dissipated.
3. The output voltage settles to the steady state level, V_{BATT}. Although not discussed here, it is possible that circulating currents may exist as a double energy transients.

Where parasitic RLC does exists, determination of the actual load circuit model may be a major problem. The designer should be aware that adding capacitance to an inductive switching circuit may create potentially damaging transients.

To determine the Energy delivered to the IC during Turn-OFF, we are only concerned with the zener diode clamp interval. As shown in the Turn-Off model of Figure 11, we have an inductance and resistance in series, connected between the clamp voltage, V_{CL} and the supply voltage, V_{BATT}. And, as noted, the clamp voltage, V_{CL}, may be treated as a voltage drive source; but only during the flyback period. When the Initial Current, I_B decays to zero, the clamp energy is depleted. The energy loss in the zener diode voltage clamp is the energy dissipated in the MOS output driver.

Given Turn-OFF boundary conditions and symbol definitions similar to Turn-ON, we have the following:

$I_L(t)$ Instantaneous Inductor Load Current where we must include the Initial Current, I_B

V_{BATT} Output Load Power Supply which is now one of two defined voltage driver sources.

V_{CL} Pulse Clamp Voltage, $V_{CL} = V_Z + V_{GS(TH)}$ and is a defined voltage drive source but only applies while $I_L(t)$ decays from the initial clamp condition at t = 0 and continues in time until $I_L(t) = 0$.

L, R_L Load Inductor, L with Resistance, R_L

T Time Constant, T = L/R where R = R_L

t_0 Turn-OFF time defined here as the active clamp period, t = 0+ to t = t_0 where $I_L = 0$ at t_0

I_O Steady State Current $I_O = \frac{V_{BATT} - V_{CL}}{R_L}$

with time limited conditions, see V_{CL} symbol definition note. Polarity is defined as positive for the V_{BATT} supply current flow to ground

I_B Initial Current $I_B = \frac{V_{BATT}}{R_L}$ at t = 0.

(or, where otherwise defined, I_B is the instantaneous current, $I_L(t)$ that is flowing at switch-off).

The direction of initial current at Turn-OFF follows that of Turn-ON conditions because current must be continuous during transient changes. The clamp voltage, V_{CL} is typically much greater than V_{BATT}, which gives an I_O (steady state) current that is negative. However, I_O does not actually reach a steady state condition because the clamp ceases to be active at t = t_0 (when $I_L(t)$ goes to zero). (At Turn-OFF, to carry the integration solution for Energy past the $I_L(t) = 0$ point will provide a fictitious answer).

To simplify the Turn-OFF calculations, energy loss is assumed to begin in the voltage clamp at t = 0 when the output driver is switched OFF. Since $I_L(t) = 0$ at t = t_0, we can set $I_L(t) = 0$ and solve for t_0 from the equation:

$$I_L(t) = I_O(1 - Ae^{-t/T}) = 0 \qquad \text{(EQ. 17)}$$

which provides the solution,

$$t_0 = T\mathrm{Ln}(A) \qquad \text{(EQ. 22)}$$

where Ln(A) is the natural log of A.

For the circuit conditions illustrated in Figure 11, we can now derive our current equation for $I_L(t)$ using Equation 17 by substituting the new boundary conditions defined for Turn-OFF. To begin, we have:

$$A = \left(1 - \frac{I_B}{I_O}\right) = \left(\frac{V_{CL}}{V_{CL} - V_{BATT}}\right) \qquad \text{(EQ. 23)}$$

The voltage ratio expression for A assumes Turn-OFF is initiated from a steady state ON condition. To avoid limitations, it is better to stay with the current ratio expression for A. As such, the output current at switch-off is equal to the instantaneous value of the ON state current calculated using Equation 17.

From Equation 21, the integral equation for Clamp Energy, E_{CLAMP} at Turn-OFF is:

$$E_{CLAMP} = V_{CL} I_O \int_0^t (1 - Ae^{-t/T})\,dt \qquad \text{(EQ. 24)}$$

When the this equation is integrated over the limits of t = 0 (at switch-OFF) to t = t, we have:

$$E_{CLAMP} = V_{CL} I_O \left(t + ATe^{-\frac{t}{T}}\right)\Bigg|_0^t \qquad \text{(EQ. 24A)}$$

Expanding Equation 24A, we have the general solution for energy loss, E_{CLAMP} in the time interval, t after Turn-OFF which is:

$$E_{CLAMP} = V_{CL} I_O T\left(\frac{t}{T} - \left(1 - Ae^{-\frac{t}{T}}\right)\right) \qquad \text{(EQ. 24B)}$$

Our primary interest is to determine the total energy loss during Turn-OFF. Given that t_0 = TLn(A), the expression for total Clamp Energy at Turn-OFF is:

$$E_{CLAMP} = V_{CL} I_O T(1 - A + \mathrm{Ln}(A)) \qquad \text{(EQ. 25)}$$

Care is needed to assure that the value for A is carried out to several decimal places because the value of A may be very close to 1, causing the (1-A) term to lose accuracy.

Example: Using component values equal to those given for Turn-ON, and a defined initial current condition for t = 5T after Turn-ON; we have for Turn-OFF:

V_{BATT} = 13.5V

V_{CL} = 82V

L = 100mH, R = R_L = 9.5Ω

$I_O = (V_{BATT} - V_{CL})/R = -7.21A$

I_B = use the calculated values from Turn-ON

for t = 5T, I_B = 1.341A (~V_{BATT}/R);

A = 1 - (1.341/(-7.21)) = 1.18596

T = L/R = 100mH/9.5Ω = 10.526ms

Then t_0 = (10.526ms) Ln(1.18596) = 1.7953ms

We can now calculate energy for the assumed example values. We should first note that t_0 is less than T, requiring a full detailed calculation without approximations. However, unlike the Turn-ON solution which goes to a steady state condition, the Turn-OFF Energy calculation is complete when t = t_0. The solution using Equation 25 is:

E_{CLAMP} = (82)(-7.2)(10.5ms)(1-1.186+Ln(1.186)) = 95.9mJ

For the examples given above, the Turn-ON and Turn-OFF current curves are plotted in Figure 12.

L and R_L Energy Calculations

At Turn-ON, the load resistance dissipates power as an energy loss while the coil inductance stores magnetic field energy. The general equation for resistive energy loss, E_R in the total series load resistance is given in Equation 16A. Since we have the same integral equation for each series component term of the loop, the Turn-ON and ON state energy loss equation for R_L is:

$$E_{RLON} = R_L I_O^2 \left(t + 2ATe^{-\frac{t}{T}} - \frac{A^2T}{2} e^{-\frac{2t}{T}} \right) \Bigg|_0^t \qquad \text{(EQ. 26)}$$

And, evaluating from t = 0 to t, we have:

$$E_{RLON} = R_L I_O^2 \left[\left(t + 2ATe^{\frac{-t}{T}} - \frac{A^2T}{2} e^{\frac{-2t}{T}} \right) - \left(2AT - \frac{A^2T}{2} \right) \right]$$

$$= R_L I_O^2 \left(t - 2AT\left(1 - e^{\frac{-t}{T}}\right) + \frac{A^2T}{2}\left(1 - e^{\frac{-2t}{T}}\right) \right) \qquad \text{(EQ. 26A)}$$

At Turn-ON, the R_{DSON} energy loss expression, E_{RDSON} and the expression for E_{RLON} are the same. Therefore, we can determine the Turn-ON energy loss for R_L by a ratiocomparison. (Refer to Equation 19).

$$E_{RLON} = E_{RDSON} \frac{R_L}{R_{DSON}} \qquad \text{(EQ. 27)}$$

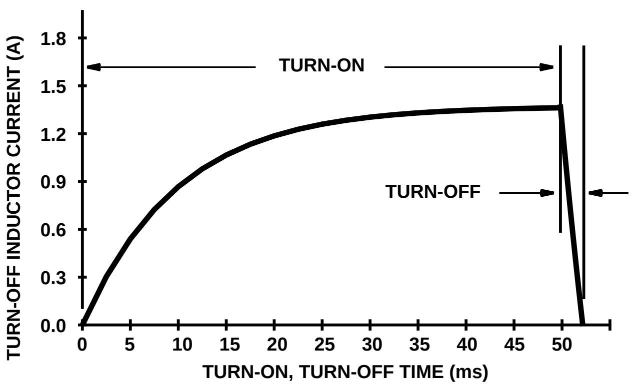

FIGURE 12A.

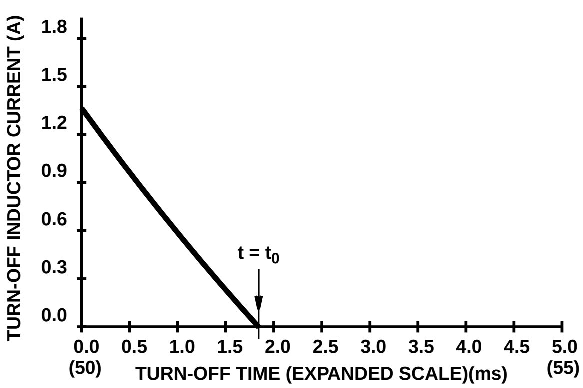

FIGURE 12B.

FIGURE 12. TRANSIENT TURN-ON AND TURN-OFF CURRENT CURVES FOR THE SWITCHED INDUCTOR LOAD EXAMPLE HAVING A TIME CONSTANT T = L/R = 10ms, V_{BATT} = 13.5V, V_{CL}= 82V, L = 100mH, R_L = 9.5Ω AND R_{DSON} = 0.5Ω

As previously noted, the Turn-ON energy loss equation continues in time as a steady state condition. Therefore, Equation 27 implies that the energy loss is over the same exact time duration.

At Turn-OFF, the period of integration is from switch off at t = 0 to t = t_0, in which case we may then substitute $t = t_0 = TLn(A)$ The solution is:

$$E_{RLOFF} = R_L I_O^2 T\left(\frac{1}{2}(A-3)(A-1) + Ln(A)\right) \qquad \text{(EQ. 28)}$$

The parameter values differ in the Turn-ON and Turn-OFF interval, as previously noted for the E_{RDSON} and E_{CLAMP} examples.

Example: Given the same conditions used for E_{CLAMP} Turn-OFF and using Equation 28, calculate the value of E_{RLOFF}.

$$E_{RLOFF} = (13.5)(-7.2)^2(10.5ms)(Ln(1.186) - 0.1687)$$

Then, $E_{RLOFF} = 9.79mJ$.

As noted in Equation 9, the stored energy in an inductor for a given current is:

$$E_L = L\left(\frac{I^2}{2}\right)$$

Using the value of I_B at t = 5T from the Turn-ON example, the stored inductive energy is:

$$E_L = 100mH\left(\frac{(1.34)^2}{2}\right) = 89.9mJ\ .$$

In the above solutions, it was assumed that the value of the coil inductance was a fixed constant value. Inductance will vary to some degree with current and temperature. The variation will dependent on the mechanical design and the quality of the ferrite material that is an integral part of the coil. Generally, as current increases, inductance will decrease. If the core material saturates, the inductance will substantially decreased.

Where the R_L load is not constant, it is possible to calculate current and energy in incremental steps. An indicator of this need would be evident in a non-exponential current rise at Turn-ON, as viewed with an oscilloscope. The inductance parameters vs current and temperature should be measured and a new time constant should be applied to each stepped increment, based on the inductance vs current data. New initial current conditions must be used for each step.

Sourced Power Supply Energy at Turn-OFF

During the discharge period, energy is sourced from the magnetic field of the inductor and from the power supply. The solution for the V_{BATT} sourced energy, E_{BATT} follows the same calculation method used in the clamp energy, E_{CLAMP} solution.

The expression for sourced energy from the battery is:

$$E_{BATT} = \int_0^{t_0} V_{BATT}\ I_L(t)\ dt \qquad \text{(EQ. 29)}$$

When integrated and evaluated over the limits, t = 0 to t = t_0 = TLn(A),

$$E_{BATT} = V_{BATT} I_O T(1 - A + Ln(A)) \qquad \text{(EQ. 29A)}$$

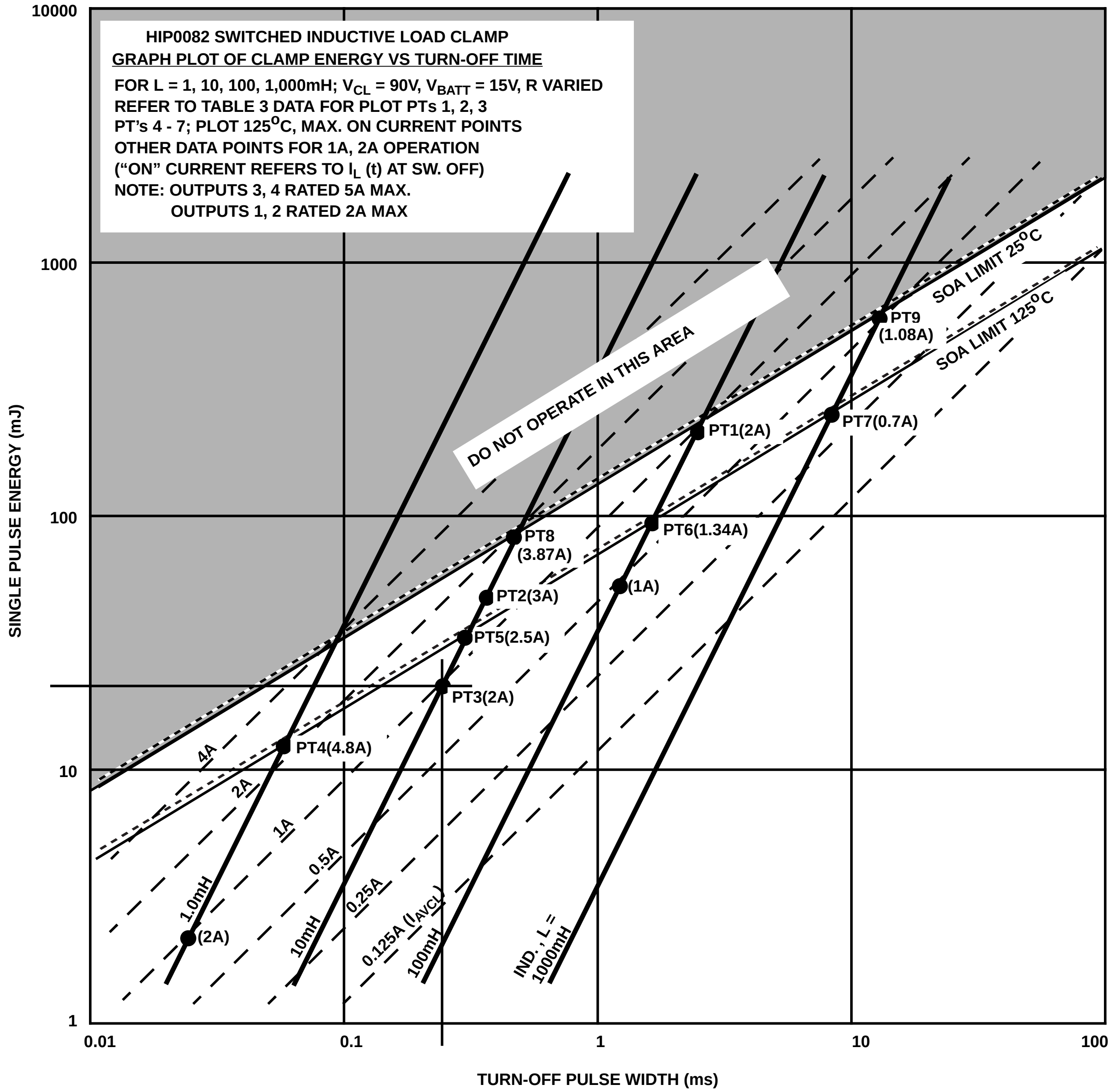

FIGURE 13. THIS SAFE OPERATING AREA (SOA) CHART FOR THE HIP0082 ILLUSTRATES THE ESSENTIAL REQUIREMENTS FOR OPERATION WITHIN THE MAXIMUM CURRENT AND ENERGY CONSTRAINTS. IN ADDITION, INITIAL OPERATING

Except for V_{BATT} in place of V_{CL}, this is the same integral solution as the one for E_{CLAMP}. While noting that E_{BATT} is sourced energy and E_{CLAMP} is lost energy, we may solve for E_{BATT} as a ratio of E_{CLAMP}.

Using ratio method, the E_{BATT} equation is:

$$E_{BATT} = E_{CLAMP}\left(\frac{V_{BATT}}{V_{CL}}\right) \qquad \text{(EQ. 30)}$$

and E_{BATT} calculates to be:

$$E_{BATT} = 95.9 \times \left(\frac{13.5}{82}\right) = 15.79\text{mJ}$$

Turn-OFF Source and Loss Energy Check

During the Turn-OFF period, the sourced power supply energy and Inductor energy should equal the energy loss in external resistor load and the voltage clamp. To check our calculations, the balanced energy equation is:

$$E_{BATT} + E_L = E_{RLOFF} + E_{CLAMP} \qquad \text{(EQ. 31)}$$

Filling in the numbers from the above calculation, we have

$$(15.8 + 89.9) = (9.8 + 95.9) = 105.7 \text{ mJ}.$$

Total IC Energy Calculation

The total energy, loss in an output driver is the sum of the Turn-ON period loss plus the Turn-OFF loss. The calculation for E_{RDSON} gave the R_{DSON} Energy loss in the output driver over Turn-ON plus ON time. For t = 0 to t = 5T in the illustrated example, the loss was 32.02mJ. During the Turn-OFF time interval, the energy loss in the IC voltage clamp,

E_{CLAMP} was calculated as 95.9mJ. For each Turn-ON and Turn-OFF cycle, the total energy loss, E_T in the output driver is:

$$E_T = (E_{RDSON} + E_{CLAMP}) = (32.02 + 95.9) = 127.9 \text{ mJ.}$$

The illustrated energy calculation based on a Turn-ON period of 5 times the L/R time constant is an arbitrary example. The user may choose any time interval for Turn-ON.

Inductive Load Switching Dissipation Calculations

Device dissipation is the energy in joules (watt-seconds) for a unit period of time. Dissipation may also be derived from the calculated energy for one switching cycle times the switching frequency, f in hertz. The Turn-ON time for t = 0 to t = 5T is 50ms and the Turn-OFF time is 1.8ms. If we assume repetitive switching with 50% ON time, the switching frequency is 10Hz. The calculated dissipation is:

$$P_D = (E_T \times f) = (127.9 \times 10) = 1279 \text{ mW.}$$

To carry the results further, assume this is an HIP0082 Quad Driver with all 4 outputs switching into an equivalent load. Then, the total dissipation is 4x1.28 = 5.12W. If we use a $6^oC/W$ external heat sink, the total $\theta_{JC} + \theta_{CA}$ is $9^oC/W$. Using Equation 12A and 150^oC for the maximum junction temperature, the solution for the maximum ambient temperature is:

$$T_A = T_J - P_D \times \theta_{JA} = (150 - 5.12 \times 9) = 104\ ^\circ C.$$

Using the Inductive Switching SOA Curve

Having defined the variables of inductive switching and equation solutions to calculate current and energy, we can now evaluate and tabulate the results. The key switching parameters for the Intersil family of Power BiMOS Low Side Drivers are given in Table 3. Using the IC specifications and a C Program (see the Appendix), current and energy data is generated for various points of operation.

Figure 13 illustrates the how R_L load current and energy data is applied to the HIP0082 Inductive Switching SOA Curve of Figure 7. For defined voltage conditions, a linear curve is plotted for different inductive load values. (The inductance curves are not intended as rating information.) The load resistance is varied to generate points on each inductive load curve.

As an example, the Table 3 data for the HIP0082 tabulates current operation at 2A in an inductive load of 10mH. Data point PT3 is a plot of the Clamp Energy, E_{CLAMP} = 21.14mJ and Turn-OFF time, t_0 = 0.243ms.

We should note as a reminder that the energy stored in the inductor ($LI^2/2$) is not equal to the clamp energy at Turn-OFF. Additional source energy is supplied from the power supply to the clamp during the Turn-OFF interval. For the examples used here, this added contribution is small. However, if the clamp voltage level is reduced, the additional source energy will increase.

For fixed values of inductance, a plot of increasing current intersects the SOA limits to define the maximum rated output drive. In Figure 13 the currents shown in parenthesis are the saturated ON levels of current at the time of switch-off. An average value of the output pulse current will more closely correlate to the plotted current lines in Figure 7. If we integrate the Turn-OFF pulse current over time t_0, the result is:

$$I_{AVCL} = I_O + I_B / Ln(A) \qquad \text{(EQ. 32)}$$

The current during the clamp interval is slightly less than one-half of I_B. The ratio of I_B to I_{AVCL} is typically in the 0.46 to 0.49 range.

It is important to note that load current at Turn-OFF (shown in parenthesis) will remain the same as long as the supply voltage, V_{BATT} and the R, L values remain the same. As such, the plotted inductance curves specifically relate to the assumed values for V_{BATT} and V_{CL}. While other parameters remain the same, a changing clamp voltage will generate different values for clamp energy, E_{CLAMP} and Turn-OFF Time, t_0. Increasing the clamp voltage will decrease the Turn-OFF time, t_0 and cause higher dissipation for shorter periods of time, reducing the SOA margin.

Sustained dissipation increases the chip junction temperature, reducing the SOA margin for safe operation. The SOA margin is reduced, as indicated by the SOA limits for 125^oC operation vs 25^oC operation.

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TABLE 3. POWER BIMOS LOW SIDE DRIVERS - FEATURES AND INDUCTIVE SWITCHING CHARACTERISTICS

PARAMETER	HIP0080 QUAD	HIP0081 QUAD	HIP0082 QUAD			CA3282 OCTAL	UNITS
Output Voltage Clamp Range Min. Max.	27 43	73 89	72 90			27 40	V
Maximum Rated DC Load Current	1	2	2A Outputs 1 and 2 5A Outputs 3 and 4			1	A
Maximum R_{DSON} Output Resistance	1.0Ω at 0.5A	0.5Ω at 1A	0.62Ω at 2A Outputs 1 and 2 0.57Ω at 2A Outputs 3 and 4			1.0Ω at 0.5A	Ω A
Minimum Output Current Limiting (Latch-OFF Threshold)	1.3	2.2	2.1A Outputs 1 and 2 5.1A Outputs 3 and 4			1.05	A
Thermal Shutdown or Flag	SD	SD	Flag			None	-
Fault, Diagnostics, Feedback	Yes	Yes	Yes			Yes	-
Plastic Package	28 PLCC	15 SIP	15 SIP			15 SIP	-
Junction-to-Case Thermal Resistance	14	3	3			3	^{o}C/W
ENERGY DATA FOR INDUCTIVE LOAD SWITCHING: Battery Supply = 15V; Clamp Voltage and R_{DSON} at Maximum Specified Value. Data shown for 5T Turn-ON Time, then Turn-OFF. (T = L/R Time Constant)							
Output Voltage Clamp (Max.), V_{CL}	43	89	90	90	90	40	V
Output Operating Current, I_O	0.5	1	2	2	3	0.5	A
Output Drain Resistance, R_{DSON}	1	0.5	0.6	0.6	0.6	1	Ω
Load Resistance, R_L	29	14.5	6.9	6.9	4.4	29	Ω
Load Inductance, L	100	100	100	10	10	100	mH
Turn-ON Time Const., $T = L/(R_L+R_{DSON})$	3.33	6.67	13.33	1.333	2	3.33	ms
Turn-OFF Time Const., $T = L/(R_L)$	3.45	6.9	14.49	1.449	2.273	3.45	ms
Turn-ON Time, t = 0 to t = 5T	16.67	33.33	66.67	6.67	10	16.67	ms
Turn-OFF Time, $t = t_0$	1.43	1.23	2.43	0.243	.366	1.57	ms
Load Current at t = 5T (Turn-OFF Initial Current Condition)	0.497	0.993	1.987	1.987	2.98	0.497	A
Average Current During Clamp Period	0.231	0.482	0.966	0.966	1.45	0.23	A
ENERGY							
IC Turn-ON Energy, t=0 to t = 5T; E_{RDSON}	2.93	11.71	112.43	11.24	37.95	2.93	mJ
Stored Inductor Energy, at t = 5T; E_L	12.33	49.23	197.31	19.73	44.4	12.33	mJ
IC Turn-OFF Clamp Energy, E_{CLAMP}	14.23	52.6	211.38	21.14	47.78	14.41	mJ
IC Turn-ON + Turn-OFF Energy	17.15	64.31	323.81	32.38	85.73	17.33	mJ
DISSIPATION							
Frequency for 50% duty cycle (on time)	30	15	7.5	75	50	30	Hz
Dissipation, Single Output Switching	515	965	2429	2429	4286	520	mW

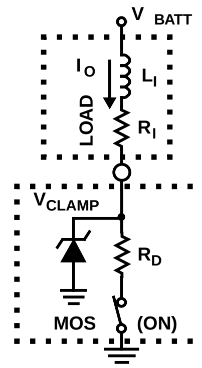
Appendix - A1: Inductive Load Calculations - C Program and Data Run

POWER BiMOS INDUCTIVE SWITCHING EXAMPLE CALCULATIONS

Input Data: HIP0082_TYPICAL_DATA
Li= 100 mH, Ri= 9.5 ohms, Vbatt= 13.5 V, Vclamp= 82 V, Rd= 0.5 ohms
****

Turn-ON: Output Load Current and IC/Rdson Energy Loss Tabulation
Initial Current is Ib=0, Steady State Current, Io=Vbatt/(Ri+Rd)
Is = Is(t) = Instantaneous Load Current
Time in 10 ms steps, t=0 to t=50 ms, T=[L/(Ri+Rd)]=10 ms

kT Steps	Time in ms	Load Is,Amps	Turn-ON Eon,mJ	Turn-ON Eon S/C	% Eon S/C Error
1	10	0.8534	1.532	-4.556	397.5
2	20	1.167	6.939	4.556	34.34
3	30	1.283	14.56	13.67	6.152
4	40	1.325	23.11	22.78	1.438
5	50	1.341	32.02	31.89	0.3829

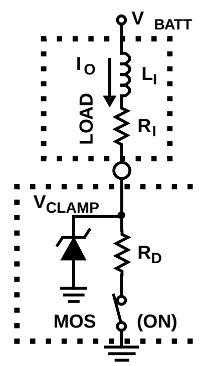


IC MOS OUTPUT DRIVER

****

Turn-OFF: Output Clamp Current and Energy Tabulation
Note: At t=to, Inductive Discharge Current depletes to Zero Amps
Initial Current, Ib = Turn-ON step at t=5T
Steady State Current, Io=(Vbatt-Vclamp)/Ri
Inc.= 5 steps from t=0 to t=to= 1.7953 ms, T=(L/Ri)= 10.53 ms

k k*to/5 Steps	Time in ms	Load Current Is,Amps	Trn-OFF Energy Ec,mJ	Delta Energy Inc.
0	0	1.34	-0	0.0
1	0.359	1.05	35.23	35.2
2	0.718	0.777	62.17	26.9
3	1.08	0.509	81.08	18.9
4	1.44	0.25	92.24	11.2
5	1.8	-0	95.9	3.66



IC MOS OUTPUT DRIVER

Average Current during Turn-OFF, Iavcl = (Io+Ib/lnA) = 0.6514
Initial Current at Switch OFF from t=5T ON State, Ib = 1.3409
Ratio: Iavcl to Ib, Initial ON Current = 0.485794

****

Application Note Example Energy Source/Loss Tabulation

Turn-OFF Sourced Energy, Eb(Batt.),Ei(L) from Switch OFF to t=to:
Sourced Battery Energy, Eb: 15.788 mJ
Stored Coil Energy at Switch-OFF, Ei: 89.901 mJ
Total Sourced Turn-OFF Energy, Eb+Ei: 105.69 mJ

Turn-OFF Energy Loss, Eclamp(IC), Er(Ri) from Switch OFF to t=to:
Clamp Energy, Eclamp: 95.898 mJ
Load Resistor Energy, Er: 9.7915 mJ
Total Loss Turn-OFF Energy, Er+Eclamp: 105.69 mJ

Total IC ON+OFF Energy loss, t=0 --> t= 5T --> t= 5T+to:
Turn-ON Energy Loss for t=0 to t= 5T, Eon: 32.016 mJ
Clamp Energy Loss at Switch Off, Eclamp: 95.898 mJ
Total IC Energy Loss, Et=Eoff(0--> 5T)+Eclamp: 127.91 mJ

Example Dissipation for ON/OFF cycle, t=0 --> t= 5T --> t= 5T+to:
Total ON/OFF Switch Time (ton+toff)=(50+ 1.8)= 51.8 ms

For Each MOS Switching Output, Dissipation = Energy x Frequency:
For repetitive switching, 50% ON time, Freq, f = 10 Hz
Single Switch Dissipation, Pd = (10 x Et) = 1279 mW

Appendix - A2: Inductive Load Calculations - C Program and Data Run (Continued)

```
/* Power BiMOS Application Note Inductive RL Switching Examples,  rev 8/11/94  */
#include <stdio.h>
#include <sys/ieeefp.h>
#include <floatingpoint.h>
#include <math.h>

main()
{
   char Name[10];
   double R,Rd,Ri,Li,T,Vb,Vc,A,Io,Ib,t,to,tson,Is,Ison;
   double x,y,Eon,Eson,Eonsc,Eoff,Ei,Eb,Ec,Ecs,Eroff;
   int k,n=5,step=1;

   printf("POWER BiMOS INDUCTIVE SWITCHING EXAMPLE CALCULATIONS\n\n");
/*    optional data entry, unix redirection from data file or type in by next 2 line prompt  */
/*    printf("Enter Data, Use only white-space or <CR> separation\n"); */
/*    printf("Input: Name Vbatt, Vclamp, Li, Ri, Rd \n"); */
   scanf("%s %lf %lf %lf %lf %lf", Name,&Vb,&Vc,&Li,&Ri,&Rd);
   printf("Input Data: %s\n", Name);
   printf("Li= %g mH, Ri= %g ohms, Vbatt= %g V, Vclamp= %g V, Rd= %g ohms\n", Li*1000, Ri, Vb, Vc,
Rd);

/* Turn ON example */
   R=Ri+Rd;
   T=Li/R;
   Io=Vb/R;
   Ib=0; /* Assumed value for this example */
   A=1-(Ib/Io);
   printf("*******************************************************************\n");
   printf("\nTurn-ON: Output Load Current and IC/Rdson Energy Loss Tabulation\n");
   printf("   Initial Current is Ib=0, Steady State Current, Io=Vbatt/(Ri+Rd)\n");
   printf("   Is = Is(t) = Instantaneous Load Current\n");
   printf("Time in %2.2g ms steps, t=0 to t=%2.3g ms, T=[L/(Ri+Rd)]=\t%2.4g ms\n\n",step*T*1000,
n*T*1000,T*1000);

   printf("  kT \t Time\t Load\tTurn-ON\tTurn-ON\t %% Eon\n");
   printf("Steps\tin ms\tIs,Amps\tEon,mJ\tEon S/C\t S/C Error\n");
   printf("-----\t-----\t------\t-------\t-------\t-------\n");
   for ( k=1; k<=n; ++k )
      {
      t=k*step*T;
      Is=Io*(1-A*exp(-t/T));
      Eon=Rd*Io*Io*((t+2*A*T*exp(-t/T)-A*A*T/2*exp(-2*t/T))-(2*A*T-A*A*T/2));
      Eonsc=(k-1.5)*T*Rd*Io*Io;
      printf(" %d\t% .4g\t% .4g\t% .4g\t% .4g\t% .4g\n", k,t*1000,Is,Eon*1000,Eonsc*1000,100*(Eon-
Eonsc)/Eon);
      if ( k == n )
         {
         Ison=Is;
         Eson=Eon;
         tson=t;
         }
      }
/* Turn OFF example */
   R=Ri;
   T=Li/R;
   Io=(Vb-Vc)/R;
   Ib=Ison; /* From Turn-ON step data, Is at n_th step  */
   A=1-(Ib/Io);
   to=T*(log(A));
/* printf("to=%g, A= %g, T= %g, Io= %g, Ib= %g, log(A)= %g\n",to,A,T,Io,Ib,log(A)); */
   printf("*****************************************************************\n");
   printf("\nTurn-OFF: Output Clamp Current and Energy Tabulation \n");
printf("    Note: At t=to, Inductive Discharge Current depletes to Zero Amps\n");
   printf("    Initial Current, Ib = Turn-ON step at t=5T\n");
   printf("    Steady State Current, Io=(Vbatt-Vclamp)/Ri\n");
```

Appendix - A3: Inductive Load Calculations - C Program and Data Run (Continued)

```
printf("Inc.= %d steps from t=0 to t=to=% .5g ms, T=(L/Ri)=\t% .4g ms\n\n", n,to*1000,T*1000);

    printf(" k \t Time\t Load \tTrn-OFF\t Delta\n");
    printf("k*to/5\t  in\tCurrent\tEnergy\t Energy\n");
    printf(" Steps\t  ms\tIs,Amps\tEc,mJ\t Inc. \n");
    printf("------\t------\t------\t------\t------\n");
    Ecs=0;
    for ( k=0; k<=5; ++k )
        {
        t=k*to/5;
        Is=(Io)*(1-A*(exp(-t/T)));
        Ec=Vc*Io*T*((t/T)-A*(1-exp(-t/T)));
        x=Ec-Ecs;
        if ( k==0 )
        printf("%d\t%2.3g\t% .3g\t  % .4g\t %s\n", k, t*1000,Is,Ec*1000," 0.0");
        if ( Is > 0 && k >= 1 )
            {
            printf("%d\t%2.3g\t% .3g\t % .4g\t % .3g\n", k, t*1000,Is,Ec*1000,x*1000);
            Ecs=Ec;
            }
        }
    Is=(Io)*(1-A*exp(-to/T)); /* to verify Is=0 at t=to */
    Eoff=(Vc)*Io*T*(1-A+log(A)); /* One time Energy Calculations for last row of Turn-OFF table */
    printf("%d\t%2.3g\t % .3g\t % .4g\t % .3g\n", k-1,to*1000,Is,Eoff*1000,(Eoff-Ecs)*1000);

    Ei=0.5*Li*Ib*Ib;
    Eb=Eoff*Vb/Vc;
    Eroff=Ri*Io*Io*T*((log(A))-(2*A)+(A*A/2)+1.5);

    printf("\nAverage Current during Turn-OFF, Iavcl = (Io+Ib/lnA) = \t%2.4g A\n", (Io+Ib/log(A)));
    printf("\tRatio: Iavcl to Ib, Initial ON Current = \t%g\n", ((T/to)+1/(1-A)));
    printf("Initial Current at Switch OFF from t=%dT ON State, Ib =\t% .5g A\n", n*step,Ison);
    printf("*****************************************************************\n");
    printf("\nApplication Note Example Energy Source/Loss Tabulation \n");

    printf("\nTurn-OFF Sourced Energy, Eb(Batt.),Ei(L) from Switch OFF to t=to:\n");
    printf("   Sourced Battery Energy, Eb:                   \t% .5g mJ\n", Eoff*Vb/Vc*1000);
    printf("   Stored Coil Energy at Switch-OFF, Ei:          \t% .5g mJ\n", Ei*1000);
    printf("   Total Sourced Turn-OFF Energy, Eb+Ei:          \t% .5g mJ\n", (Eb+Ei)*1000);

    printf("\nTurn-OFF Energy Loss, Eclamp(IC), Er(Ri) from Switch OFF to t=to:\n");
    printf("     Clamp Energy, Eclamp:                         \t% .5g mJ\n", Eoff*1000);
    printf("     Load Resistor Energy, Er:                     \t% .5g mJ\n", Eroff*1000);
    printf("     Total Loss Turn-OFF Energy, Er+Eclamp:         \t% .5g mJ\n", (Eroff+Eoff)*1000);

    printf("\nTotal IC ON+OFF Energy loss, t=0 --> t=%dT --> t=%dT+to: \n", n*step,n*step);
    printf("     Turn-ON Energy Loss for t=0 to t=%dT, Eon:     \t% .5g mJ\n", n*step,Eson*1000);
    printf("     Clamp Energy Loss at Switch Off, Eclamp:      \t% .5g mJ\n", Eoff*1000);
    printf("     Total IC Energy Loss, Et=Eoff(0-->%dT)+Eclamp: \t% .5g mJ\n",
n*step,(Eson+Eoff)*1000);

    printf("\nExample Dissipation for ON/OFF cycle, t=0 --> t=%dT --> t=%dT+to:\n",n*step,n*step);
    printf("     Total ON/OFF Switch Time (ton+toff)=(% .3g+% .3g)= %2.4g ms\n", tson*1000, to*1000,
(tson+to)*1000);
    printf("\nFor Each MOS Switching Output, Dissipation = Energy x Frequency:\n");
    printf("     For repetitive switching, 50%% ON time, Freq, f = \t %2.3g Hz\n", 1/(2*tson));
    printf("     Single Switch Dissipation, Pd = (%2.3g x Et) = \t%2.4g mW\n\n", (1/(2*tson)),(1/
(2*tson))*(Eson+Eoff)*1000);
}
```