Linear and Switching Voltage Regulator Fundamentals

Abstract

This paper will enable the user to understand the operation of switching and linear voltage regulators. The most commonly used regulating modes will be covered.

For linear regulators, the Standard, Low-Dropout, and Quasi Low-Dropout regulators will be covered (along with circuit examples).

In the switching regulator section, the Buck, Buck-boost, Boost, and Flyback topologies will be detailed. Some examples will be given of products available for the design and implementation of switching converters.

LINEAR VOLTAGE REGULATORS

Introduction

The linear regulator is the basic building block of nearly every power supply used in electronics. The IC linear regulator is so easy to use that it is virtually foolproof, and so inexpensive that it is usually one of the cheapest components in an electronic assembly.

This paper will present information that gives the user greater understanding of how a linear regulator works, and will help to de-mystify regulator specifications and applications.

Some typical circuits will be presented to highlight the commercial regulators that are currently available. The primary focus of the new product examples is in the area of Low-dropout regulators, which offer great advantages over standard regulators in many applications.

Linear Voltage Regulator Operation

Introduction

Every electronic circuit is designed to operate off of some supply voltage, which is usually assumed to be constant. A voltage regulator provides this constant DC output voltage and contains circuitry that continuously holds the output voltage at the design value regardless of changes in load current or input voltage (this assumes that the load current and input voltage are within the specified operating range for the part).

The Basic Linear Regulator

A linear regulator operates by using a voltage-controlled current source to force a fixed voltage to appear at the regulator output terminal (see Figure 1).

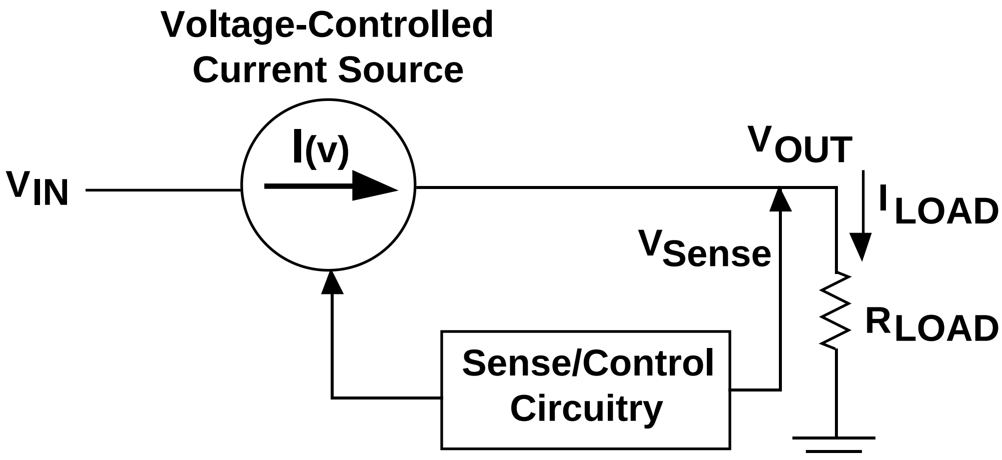


FIGURE 1. LINEAR REGULATOR FUNCTIONAL DIAGRAM

The control circuitry must monitor (sense) the output voltage, and adjust the current source (as required by the load) to hold the output voltage at the desired value. The design limit of the current source defines the maximum load current the regulator can source and still maintain regulation.

The output voltage is controlled using a feedback loop, which requires some type of compensation to assure loop stability. Most linear regulators have built-in compensation, and are completely stable without external components. Some regulators (like Low-Dropout types), do require some external capacitance connected from the output lead to ground to assure regulator stability.

Another characteristic of **any** linear regulator is that it requires a finite amount of time to "correct" the output voltage after a change in load current demand. This "time lag" defines the characteristic called **transient response**, which is a measure of how fast the regulator returns to steady-state conditions after a load change.

Control Loop Operation

The operation of the control loop in a typical linear regulator will be detailed using the simplified schematic diagram in Figure 2 (the function of the control loop is similar in all of the linear regulator types).

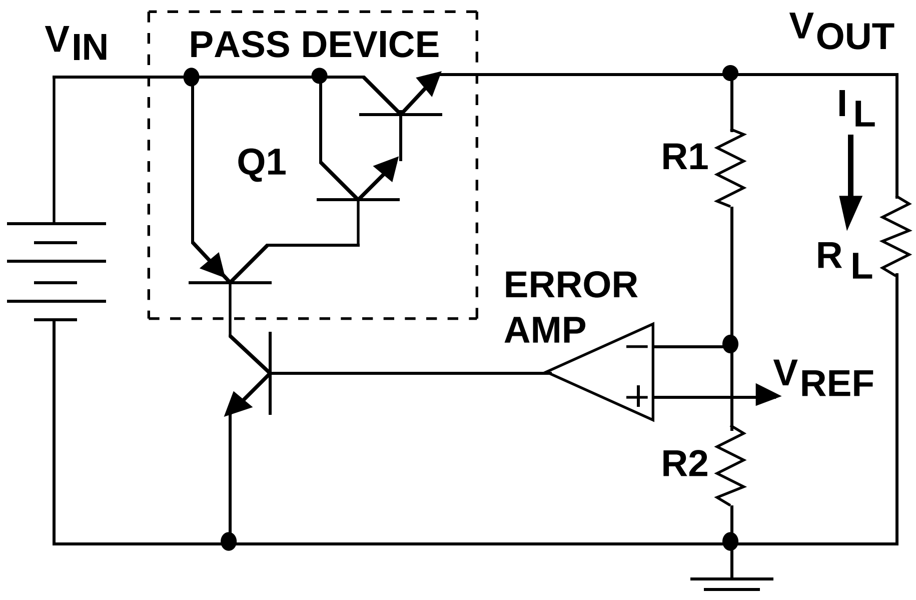


FIGURE 2. DIAGRAM OF A TYPICAL LINEAR REGULATOR

The **pass device (Q1)** in this regulator is made up of an NPN Darlington driven by a PNP transistor (this topology is a **Standard** regulator, as detailed in the following section). The current flowing out the emitter of the pass transistor (which is also the load current I_L) is controlled by Q2 and the voltage error amplifier. The current through the R1, R2 resistive divider is assumed to be negligible compared to the load current.

The feedback loop which controls the output voltage is obtained by using R1 and R2 to "sense" the output voltage, and applying this sensed voltage to the inverting input of the voltage error amplifier. The non-inverting input is tied to a reference voltage, which means the error amplifier will constantly adjust its output voltage (and the current through Q1) to force the voltages at its inputs to be equal.

The feedback loop action continuously holds the regulated output at a fixed value which is a multiple of the reference voltage (as set by R1 and R2), regardless of changes in load current.

It is important to note that a **sudden** increase or decrease in load current demand (a "step" change in load resistance) will cause the output voltage to change until the loop can correct and stabilize to the new level (this is called **transient response**). The output voltage change is sensed through R1 and R2 and appears as an "error signal" at the input of the error amplifier, causing it to correct the current through Q1.

Linear Regulator Types (LDO, Standard, and Quasi-LDO)

There are three basic types of linear regulator designs which will be covered:

Standard (NPN Darlington) Regulator

Low Dropout or LDO Regulator

Quasi LDO Regulator

The single most important difference between these three types is the **dropout voltage**, which is defined as the **minimum voltage drop required across the regulator to maintain output voltage regulation**. A critical point to be considered is that the linear regulator that operates with the smallest voltage across it **dissipates the least internal power and has the highest efficiency**. The **LDO** requires the **least** voltage across it, while the **Standard** regulator requires the **most**.

The second important difference between the regulator types is the **ground pin current** required by the regulator when driving rated load current. The **Standard regulator has the lowest ground pin current**, while the **LDO generally has the highest** (differences between the types is detailed in the following sections). Increased ground pin current is undesirable since it is "wasted" current, in that it must be supplied by the source but does not power the load.

THE STANDARD (NPN) REGULATOR

The first IC voltage regulators made used the NPN Darlington configuration for the pass device, and are designated as the **Standard** regulator (see Figure 3).

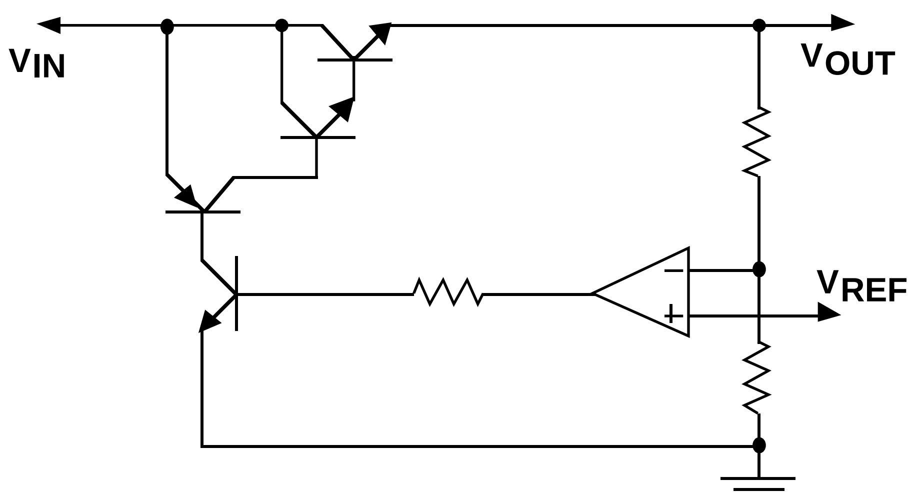


FIGURE 3. STANDARD (NPN) REGULATOR

An important consideration of the Standard regulator is that to maintain output regulation, the pass transistor requires a **minimum** voltage across it given by:

$$V_{D(MIN)} = 2\,V_{BE} + V_{CE} \qquad \text{(Standard Regulator)}$$

Allowing for the -55°C to +150°C temperature range, this minimum voltage requirement is usually set at about **2.5V to 3V** by the manufacturer to guarantee specified performance limits.

The voltage where the output actually falls out of regulation (called the **dropout voltage**) will probably be somewhere between 1.5V and 2.2V for a Standard regulator (it is **dependent on both load current and temperature**). **The dropout voltage of the Standard regulator is the highest (worst)** of the three types.

The ground pin current of the Standard regulator is very low (an LM309 can supply 1A of load current with less than 10 mA of ground pin current). The reason for this is that the **base drive current to the pass transistor (which flows out the ground pin) is equal to the load current divided by the gain of the pass device**. In the Standard regulator, the pass device is a network composed of one PNP and two NPN transistors, which means the total current gain is extremely high (>300).

The result of using a pass device with such high current gain is that very little current is needed to drive the base of the pass transistor, which results in less ground pin current. **The ground pin current of the Standard regulator is the lowest (best)** of the three regulator types.

THE LOW-DROPOUT (LDO) REGULATOR

The Low-dropout (LDO) regulator differs from the Standard regulator in that the pass device of the LDO is made up of only a single PNP transistor (see Figure 4).

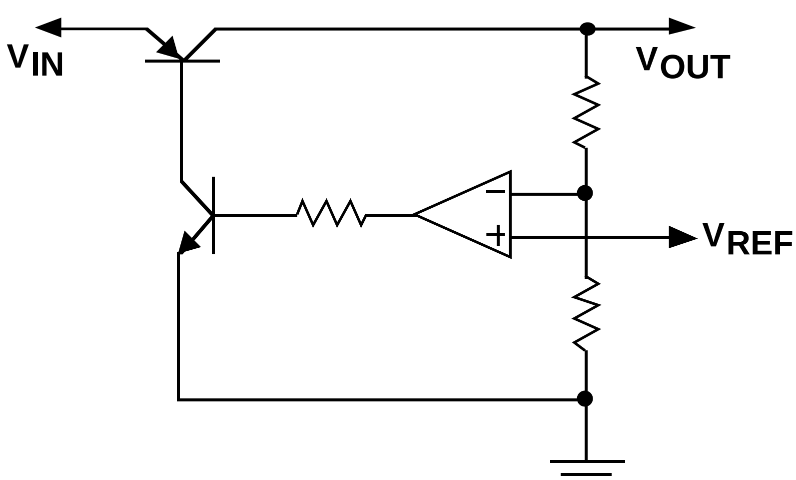


FIGURE 4. LDO REGULATOR

The **minimum** voltage drop required across the LDO regulator to maintain regulation is just the voltage across the PNP transistor:

$$V_{D(MIN)} = V_{CE} \qquad \text{(LDO Regulator)}$$

The **maximum specified dropout voltage of an LDO regulator is usually about 0.7V to 0.8V at full current**, with typical values around 0.6V. The dropout voltage is directly related to load current, which means that at very low values of load current the dropout voltage may be as little as 50 mV. The **LDO regulator has the lowest (best) dropout voltage specification of the three regulator types**.

The lower dropout voltage is the reason LDO regulators dominate battery-powered applications, since they maximize the utilization of the available input voltage and can operate with higher efficiency. The explosive growth of battery-powered consumer products in recent years has driven development in the LDO regulator product line.

The ground pin current in an LDO regulator is approximately **equal to the load current divided by the gain of the single PNP transistor**. Consequently, the **ground pin current of an LDO is the highest of the three types.**

For example, an **LP2953** LDO regulator delivering its full rated current of 250 mA is specified to have a ground pin current of 28 mA (or less), which translates to a PNP gain of 9 or higher. The **LM2940** (which is a 1A LDO regulator) has a ground pin current specification of 45 mA (max) at full current. This requires a current gain of no less than 22 for the PNP pass transistor at rated current.

THE QUASI LOW-DROPOUT REGULATOR

A variation of the Standard regulator is the quasi-LDO, which uses an NPN and PNP transistor as the pass device (see Figure 5):

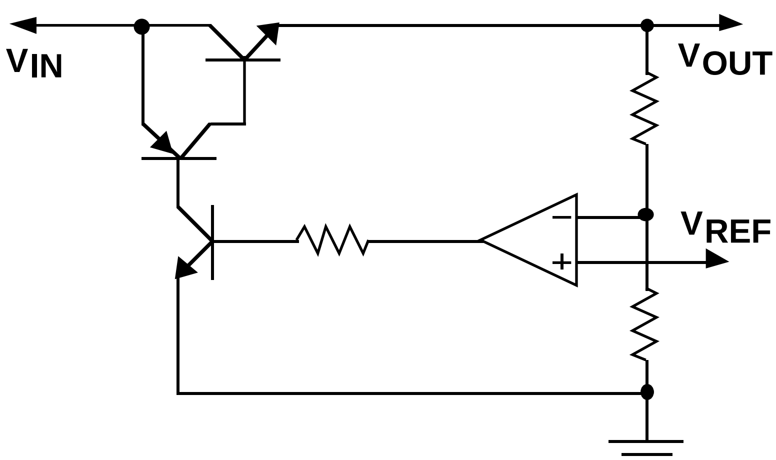


FIGURE 5. QUASI-LDO REGULATOR

The **minimum** voltage drop required across the Quasi-LDO regulator to maintain regulation is given by:

$$V_{D(MIN)} = V_{BE} + V_{CE} \qquad \text{(QUASI-LDO Regulator)}$$

The dropout voltage for a quasi-LDO delivering rated current is usually specified at about **1.5V(max)**. The actual dropout voltage is temperature and load current dependent, but could never be expected to go lower than about 0.9V (25°C) at even the lightest load. **The dropout voltage for the quasi-LDO is higher than the LDO, but lower than the Standard regulator**.

The ground pin current of the quasi-LDO is fairly low (usually less than 10mA for full rated current) which is as good as the Standard regulator.

SUMMARY

A comparison of the three regulator types[1] is shown in Figure 6.

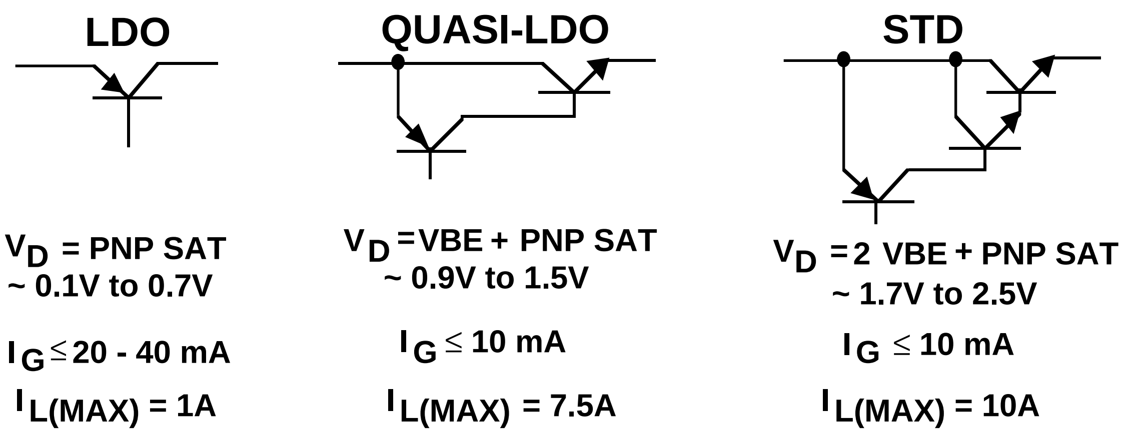


FIGURE 6. REGULATOR TYPE COMPARISON

The Standard regulator is usually best for AC-powered applications, where the low cost and high load current make it the ideal choice. In AC-powered applications, the voltage across the regulator is usually at least 3V or more, so dropout voltage is not critical.

Interestingly, in this type of application (where the voltage drop across the regulator is > 3V) Standard regulators are actually more efficient than LDO types (because the Standard has much less internal power dissipation due to ground pin current).

The LDO regulator is best suited for battery-powered applications, because the lower dropout voltage translates directly into cost savings by reducing the number of battery cells required to provide a regulated output voltage. If the input-output voltage differential is low (like 1V to 2V) the LDO is more efficient than a Standard regulator because of reduced power dissipation resulting from the load current multiplied times the input-output voltage differential.

1. The designations Standard, LDO, and Quasi-LDO **as used in this paper are by no means uniform practice throughout the industry**. At this time, National Semiconductor makes only the Standard NPN and LDO regulators (no quasi-LDO types), which means all of our LDO regulators use the single PNP device pass transistor and have dropout voltages < 1V. Another major manufacturer makes only the Standard and quasi-LDO regulators, **but advertises and sells the quasi-LDO parts as "Low Dropout Regulators"**. Still another manufacturer (who makes both the quasi-LDO and LDO types) sells the **quasi-LDO units as "Low Dropout" and the LDO units as "Very Low Dropout"**. It is strongly recommended that the designer read the fine print on the data sheet, to find out what the part will actually do **(and not rely on advertising descriptions)**.

Selecting the Best Regulator For Your Application

The best choice for a specific application can be determined by evaluating the requirements such as:

Maximum Load Current

Type of Input Voltage Source (Battery or AC)

Output Voltage Precision (Tolerance)

Quiescent (Idling) Current

Special Features (Shutdown Pin, Error Flag, etc.)

MAXIMUM LOAD CURRENT

The maximum current required in an application should be carefully considered when selecting an IC regulator. The load current specification for an IC regulator will be defined as either a **single value** or a value that is **dependent on input-output voltage differential** (this is detailed in the following section on protection circuits).

The regulator selected must be able to provide sufficient current to the load under worst-case operating conditions, if system performance is to be reliable.

INPUT VOLTAGE SOURCE (BATTERY OR AC)

The available input voltage (battery or AC power) will strongly influence which type of regulator is best suited for an application.

Battery: In battery-powered applications, **LDO regulators are usually the best choice** because they utilize the available input voltage more fully (and can operate longer into the discharge cycle of the battery).

For example, a "6V' lead-acid battery (a popular battery type) has a terminal voltage of about 6.3V when fully charged, and about 5.5V at the end-of-discharge point. If a designer wanted to make a regulated 5V supply powered from this battery, **an LDO regulator would be required** (because there is only about 0.5V to 1.3V available for dropout voltage).

AC: If a DC supply is generated from a rectified AC source, the dropout voltage of the regulator is not as critical because additional regulator input voltage is easily obtained by increasing the secondary voltage of the AC transformer (by adding turns to the secondary winding).

In these applications, a standard regulator is **usually** the most economical choice and can also provide more load current. However, in some cases the **additional features** and **better output voltage precision** of some of the new LDO regulators would still make them the best choice.

OUTPUT VOLTAGE PRECISION (TOLERANCE)

Typical linear regulators usually have an output voltage specification that guarantees the regulated output will be within 5% of nominal. This level of accuracy is adequate for most applications.

There are many new regulators which have tighter output tolerances (better than 2% is common), achieved through the use of a laser-trim process. Also, many of the new regulators have separate output specifications that cover room temperature/full operating temperature range, and full-load/light-load conditions.

QUIESCENT (IDLING) CURRENT

The quiescent current that a part draws from the source when idling (either shut down or not delivering significant amounts of load current) can be of critical importance in battery-powered applications.

In some applications, a regulator may spend most of its life shut off (in standby mode) and only supply load current when a main regulator fails. In these cases, the quiescent current determines the battery life.

Many of the new LDO regulators are optimized for low quiescent current (like 75 to 150 μA), and provide significant improvement over typical regulators which draw several milliamps.

SPECIAL FEATURES

Many LDO regulators offer features that allow the designer greater flexibility:

Shutdown: A low-power shutdown pin allows a regulator to be switched off by a logic gate or microcontroller. This feature also allows wiring a regulator for "Snap-ON/Snap-OFF" operation, which will be covered in one of the design examples presented later.

Load-dump Protection: Regulators used in automotive applications need built-in protection against overvoltage transients (load-dump). In these cases the regulator usually shuts down its output during the overvoltage transient, then recovers after it has passed.

Reverse Input Voltage Protection: This prevents damage to the regulator when the input voltage is reversed, essential in applications where the user can accidentally reverse the polarity of the batteries.

Error Flag: This flag is used to alert monitoring or control circuitry that the output has dropped about 5% below its nominal value. It is intended as a "warning flag" that can alert a controller that supply voltage may be low enough to cause erratic operation of the CPU or associated logic circuits.

Protection Circuits Built Into IC Linear Regulators

Linear IC regulators contain built-in protection circuits which make them virtually immune to damage from either excessive load current or high operating temperature. The two protection circuits found in nearly all linear IC regulators are:

Thermal Shutdown

Current Limiting

CHAIN OF COMMAND

The thermal shutdown, current limiter, and voltage error amplifier make up three distinct and separate control loops that have a definite hierarchy (pecking order) which allows one to "override" the other. The order of command (and importance) of the loops is:

1) Thermal Limit (IC is regulating junction temperature/power dissipation)
2) Current Limit (IC is regulating load current)
3) Voltage Control (IC is regulating output voltage)

This hierarchy means that a linear regulator will normally try to operate in "constant voltage" mode, where the voltage error amplifier is regulating the output voltage to a fixed value. However, **this assumes that both the load current and junction temperature are below their limit threshold values**.

If the load current increases to the limiting value, the current limiting circuitry will take control and force the load current to the set limiting value (overriding the voltage error amplifier). The voltage error amplifier can resume control only if the load current is reduced sufficiently to cause the current limiting circuits to release control. This is covered in detail in the "Current Limiting" section.

A rise in die temperature (regardless of cause) approaching the limit threshold (about 160°C) will cause the thermal shutdown to cut drive to the power transistor, thereby reducing load current and internal power dissipation. Note that the thermal limiter can override **both** the current limit circuits and the voltage error amplifier. Thermal shutdown is detailed in the next section.

It is important to understand that a **regulator holds its output voltage fixed only when it is in constant voltage mode**. In **current limiting, the output voltage will be reduced** as required to hold the load current at the set limiting value.

In thermal limiting, the output voltage drops and the load current can be reduced to any value (including zero). **No performance characteristic specifications apply when a part is operating in thermal shutdown mode**.

THERMAL SHUTDOWN

The thermal shutdown circuitry in an IC prevents the junction temperature from rising high enough to damage the part (see Figure 7). This is accomplished by monitoring the die temperature and reducing internal power dissipation to hold the temperature at the limiting value (usually about 160°C).

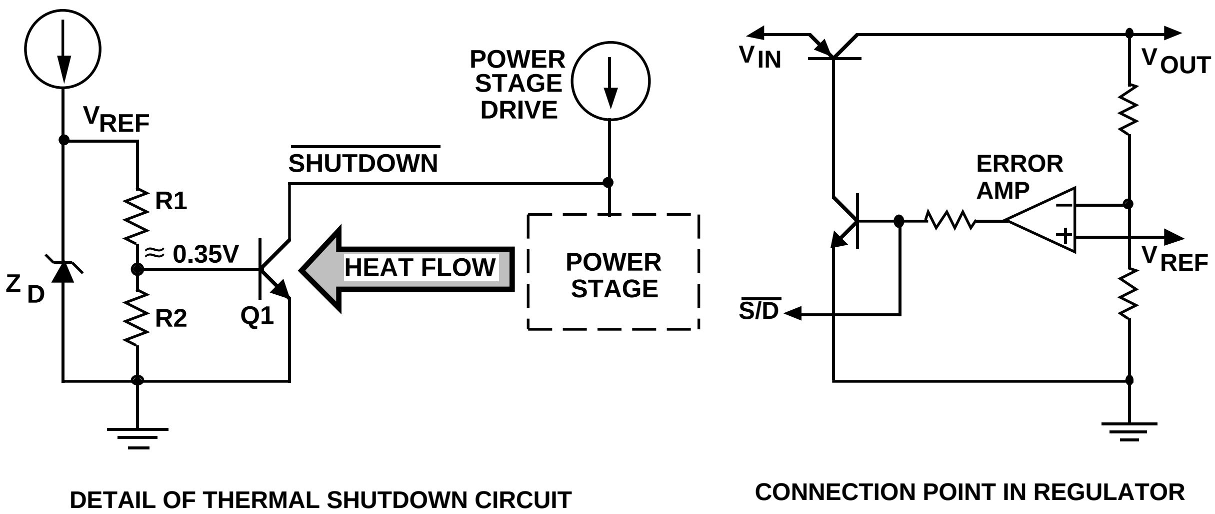


FIGURE 7. THERMAL SHUTDOWN

<u>Circuit Operation</u>:

The temperature sensor (Q1) is located near the power transistor on the die, to assure very close thermal tracking. R1 and R2 hold the base of Q1 at about 0.35V, which corresponds to the turn-on V_{BE} of Q1 at a temperature of about 160°C.

As the die temperature increases, Q1 eventually reaches the turn-on threshold (about 160°C), and starts pulling current away from the current source which supplies drive to the power stage. In this way, the load current is reduced (or cut off entirely) which reduces the internal power dissipation of the regulator.

In cases where thermal limiting occurs, both the output voltage and current will be reduced. When the output voltage drops below its nominal value, the error signal appearing at the voltage error amplifier will cause it to try and correct the regulator output voltage by driving its output high (and sourcing more current to the pass transistor).

The thermal limit circuit can sink all of the current from the error amplifier output, and keep the regulator output voltage/current as low as needed to maintain the junction temperature at 160°C. As shown, the thermal limiter can "override" the voltage control loop when needed to prevent damage to the IC.

CURRENT LIMITING

The function of current limiting circuitry is to prevent damage to the IC when an overload is placed on the output of the regulator (the load impedance is too low). Without current limiting, the regulator would source excessive load current and destroy the pass transistor inside the part.

To prevent this occurrence, the current limit circuit will override the voltage control loop, and cut down the drive to the pass transistor so that the maximum safe current level is not exceeded .

There are two basic types of current limiting circuits most commonly used in linear regulators (detailed in the next sections):

Constant Current Limiting

Voltage-Dependent Current Limiting (sometimes called "Foldback Limiting")

CONSTANT CURRENT LIMITING

The maximum current that a linear regulator can supply to a load is specified on the data sheet. Many regulators (and most LDO regulators) specify only a single value of maximum current. This value is guaranteed for any input/output voltage within the maximum ratings for the part.

For example, the LP2952 is guaranteed to source **at least 250 mA** without going into current limiting, as long as the output is in the 1.25V - 29V range and the input voltage is at least 0.8V above the output.

In Figure 8, a simplified schematic diagram is shown of a circuit that will provide constant current limiting. This is a "discrete" design implementation (the circuitry used in an IC regulator may be slightly different).

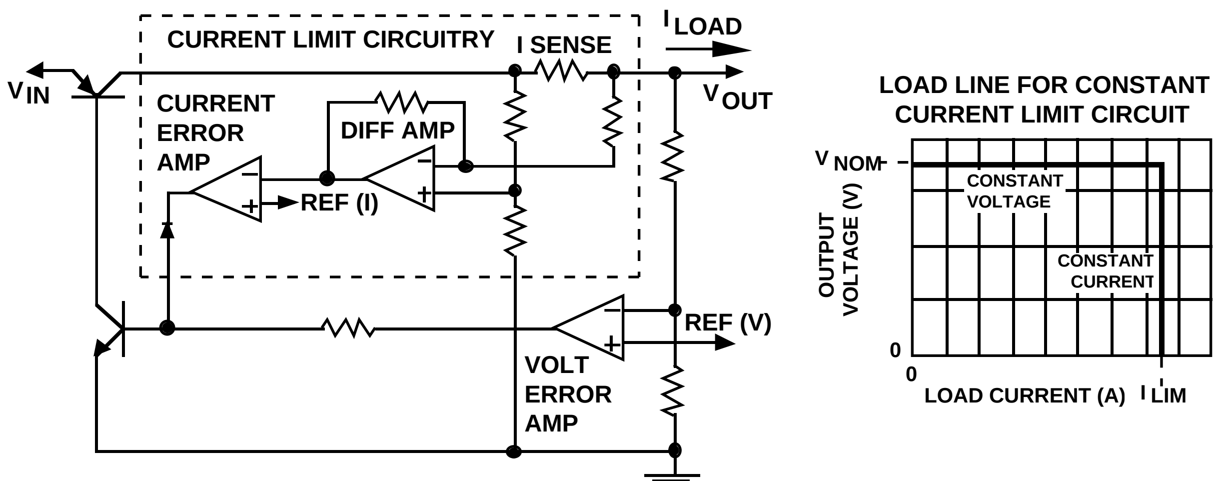


FIGURE 8. CONSTANT CURRENT LIMIT CIRCUIT

Circuit Operation:

The load current is sensed by the "I SENSE" resistor, which develops a voltage that is directly related to the current. This voltage is level shifted (and amplified) by the differential amplifier.

The voltage at the output of the differential amplifier is a ground-referenced signal that is proportional to the load current. This "load current" signal coming from the differential amplifier is applied to the inverting input of the current limit error amplifier, while the non-inverting input is connected to a reference voltage. The value of this reference voltage would be equal to the voltage at the output of the differential amplifier when the regulator is driving maximum current (at the current limit point).

Note that as long as the load current is below the limit threshold, the output of the current error amplifier is high (and the voltage error amplifier keeps the regulator in **constant voltage** mode).

When the load current reaches the limit threshold, the output of the current error amplifier drops low and starts sinking current away from the output of the voltage error amplifier (this puts the regulator in **constant current** mode).

When current limiting occurs, the regulator output voltage will drop below its nominal value, which will be sensed by the voltage error amplifier as an undervoltage condition. The voltage error amplifier will drive its output high in an attempt to raise the output voltage, but the current error amplifier can sink all of the current coming from the voltage error amplifier. Like the thermal limiter, the current limiter overrides the voltage error amplifier to prevent damage to the IC.

The load line shown in Figure 8 illustrates how the output voltage is held constant up to the point where the load current reaches the limit value, where the regulator crosses over into constant current mode. **When operating in constant current mode**, the IC regulates the load current to the "limit" value, which means **the output voltage may be any value down to zero volts**.

It should be made clear that the thermal limiter can always override the current limiter, and **can reduce the output voltage and current to any value necessary to maintain a junction temperature of about 160°C**.

For example, if the **LP2952** (which is rated for **250 mA**) is shorted from the output to ground, a **current will flow from the output which is greater than 250 mA but less than 530 mA** (see "Current Limit specification on the data sheet).

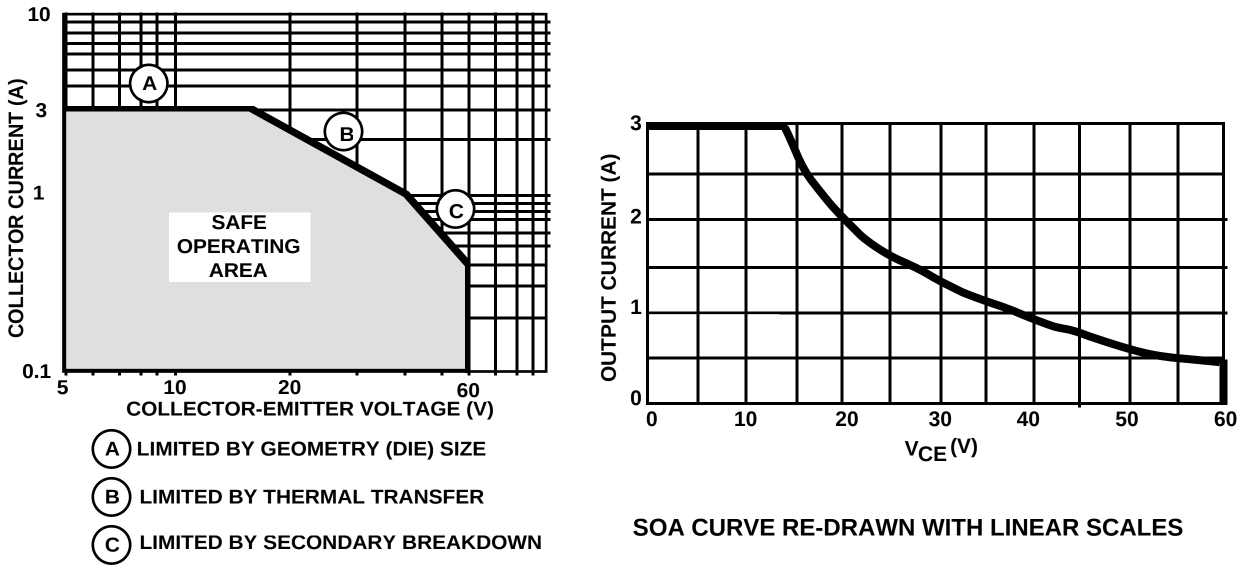
However, if the input voltage is high enough to generate sufficient power to activate the thermal limiter, that current will drop off as the LP2952 regulates its die temperature to about 160°C.

Important: Current limit circuits are (by necessity) very high-speed circuits, and input bypass capacitors on the regulator are always recommended to prevent possible device failure due to interaction with the input source impedance.

VOLTAGE DEPENDENT (FOLDBACK) CURRENT LIMITING

Voltage regulators which are relatively high current (>1A) use a type of current limiting where the maximum allowable value of load current is dependent on the input-output voltage differential across the part.

The reason this is required is due to a characteristic of all transistors called Safe Operating Area (SOA) that limits the amount of current a transistor can safely handle as the voltage increases (see Figure 9).



SOA CURVE FOR 3A/60V NPN TRANSISTOR

SOA CURVE RE-DRAWN WITH LINEAR SCALES

FIGURE 9. SOA CURVES FOR 3A/60V NPN TRANSISTOR

The data shown in the SOA curve were taken from a published data sheet for a TIP31A (3A/60V) NPN transistor. The important information on the SOA curve is that the safe **operating current value drops to about 15% of maximum when the voltage across the part (V_{CE}) is at its full rated value**. If the full **3A current rating is to be used, the V_{CE} can not exceed about 14V**.

It is important to realize that **the input-output voltage across a linear regulator is also the V_{CE} across its pass transistor**. This means **the load current must be limited in accordance with the SOA curve of the regulator pass transistor**.

The current limit curve for a linear regulator must fit **"under"** the SOA curve for the pass transistor if the device is to survive under all overload conditions. The current limit curve for the LM317 will be detailed later (in Figure 11) to illustrate this. It can be seen the shape of the curve resembles the SOA curve in Figure 9 drawn on linear axes.

CONSTANT CURRENT vs. FOLDBACK LIMITING

Constant current and foldback limiting have different characteristics that have the potential to cause some confusion.

Assuming that the designer wished to test the current limiting, he could use an adjustable power resistor connected to the output of the regulator (see Figure 10). As the resistance is adjusted to lower values (and the load current increases), the point will eventually be reached where current limiting occurs.

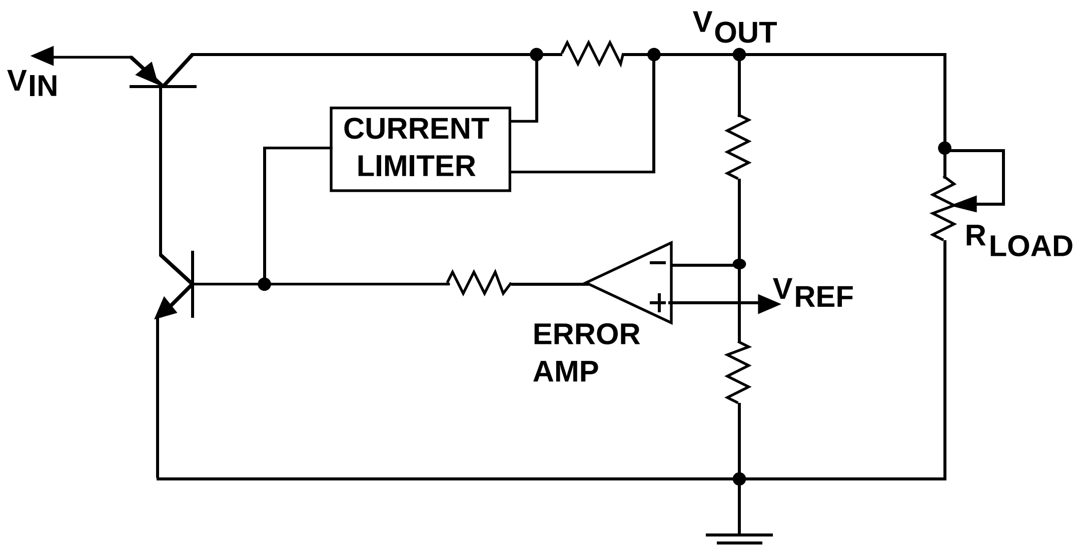


FIGURE 10. CURRENT LIMIT TEST CIRCUIT

<u>**Constant Current Limiting**</u>**:** When current limiting first occurs, the output voltage is seen to drop from its nominal value as the regulator goes from constant voltage mode to constant current mode of operation.

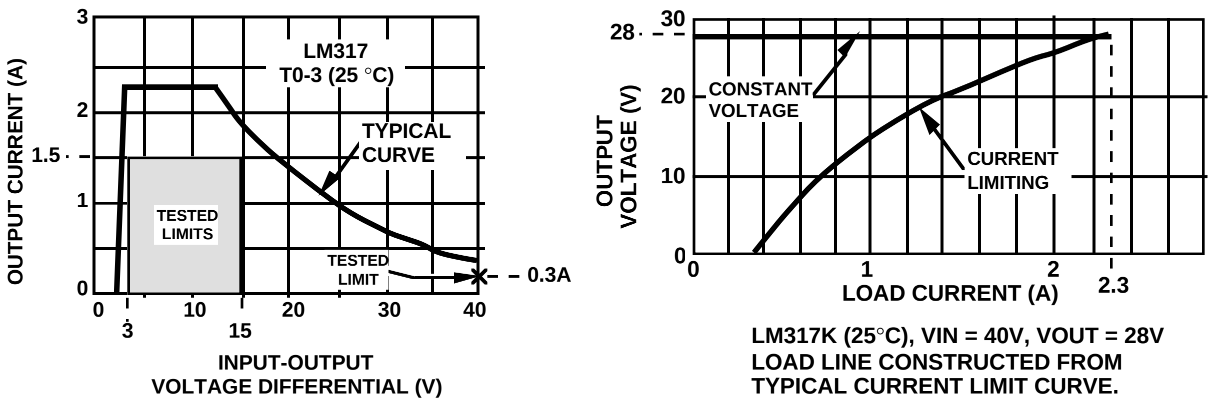
As the load resistance is decreased and current limiting occurs, **the amount that the output voltage drops is directly proportional to the decrease in load resistance** (because the load current is held **constant**).

The drop in output voltage can be made to occur gradually, and the output voltage can be moved up and down by adjusting the load resistance.

If the load resistance is increased above the point where the current limiter activated, the regulator will automatically go back into constant voltage mode (the output voltage will be in regulation).

<u>Foldback Limiting</u>: The action of a foldback limiting circuit is different, because it has some **"hysteresis"** built in to it. As the load resistance decreases to the point where limiting occurs, the output voltage can drop suddenly from the nominal voltage to a much lower value.

Returning the load resistance back to the value where limiting started may not restore the output voltage to nominal (the load resistance may have to be increased to a **higher value** to allow the regulator to return to constant voltage operation). This apparent "hysteresis" is due to the shape of the "foldback" current limit curve (see Figure 11).



LM317 CURRENT LIMIT CURVE **CURRENT LIMIT LOAD LINE**

FIGURE 11. FOLDBACK CURRENT LIMIT EXAMPLE

The example shown by the load line was constructed using the typical current limit values, assuming V_{IN} = 40V and V_{OUT} = 28V. The shape of the load line explains why the term **"foldback"** is applied, as the load current values are seen to drop with decreasing output voltage.

Explaining how foldback limiting causes hysteresis requires presenting the information in Figure 11 in a slightly different way:

The portion of the load line showing current limiting will be used to generate data points of load resistance that are equivalent to each voltage/current value along the curve (the constant voltage portion is not plotted).

The current limit resistance load line (shown in Figure 12) represents the load resistance values which correspond to the various operating points while the regulator is in the current limiting region of operation.

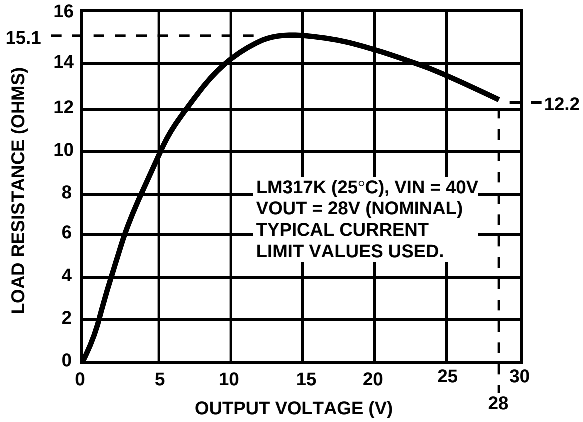
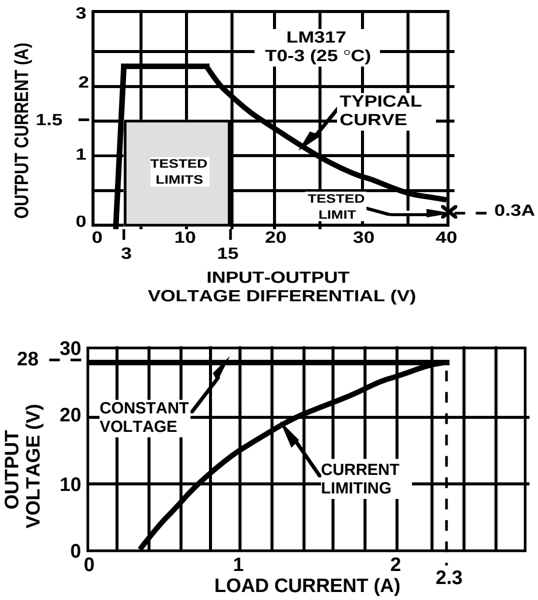


FIGURE 12. RESISTANCE LOAD LINE FOR LM317 EXAMPLE (IN CURRENT LIMITING)



LM317K (25°C), VIN = 40V, VOUT = 28V
LOAD LINE CONSTRUCTED FROM TYPICAL CURRENT LIMIT CURVE.

The curve in Figure 12 can explain how foldback limiting can cause a "hysteresis" in the output voltage as the load resistance is decreased. For example:

1) Assume the **output is at 28V** (constant voltage operation) and the **load resistance is set to 14Ω (I_L = 2A)**. The load resistance is then gradually **reduced to 12.2Ω**. The load current will then be sufficient to cause current limiting (since this is the value shown on the curve for V_O = 28V), and the output voltage will abruptly drop to the point on the load line equal to 12.2Ω. **This point corresponds to an output voltage of about 7V**.

2) In an attempt to restore the output to constant voltage operation (V_O = 28V), the load resistance is returned to 14Ω, where it had been operating previously with a 28V output. **Doing this will not return the output to 28V**, rather the operating point will go back up the load line to the first point where 14Ω is seen (at **V_O = 10V**).

To get the output back up to 28V, the load resistance has to be increased above 15.1Ω, so the operating point can get "over the bump" in the curve. If the resistance were increased gradually, the output voltage would climb slowly up to about 14V and then "jump" up to 28V.

With the example shown, **there is no value of load resistance that can be placed on the regulator output to force it to operate at output voltages between 14V and 28V**. This is the cause of the "hysteresis" that can be seen in some applications where a regulator with foldback limiting is operated at a load current where the limiting action can be made to occur.

Application Hints for Linear Regulators

Application information will be presented on subjects related to mistakes often made in applying linear regulators.

Output Capacitance Affecting Regulator Loop Stability

The output capacitor used on an LDO linear regulator can make it oscillate if the capacitor is not selected correctly.

CAPACITOR PARASITICS

Every real capacitor contains unwanted parasitic elements which degrade its electrical performance (see Figure 13).

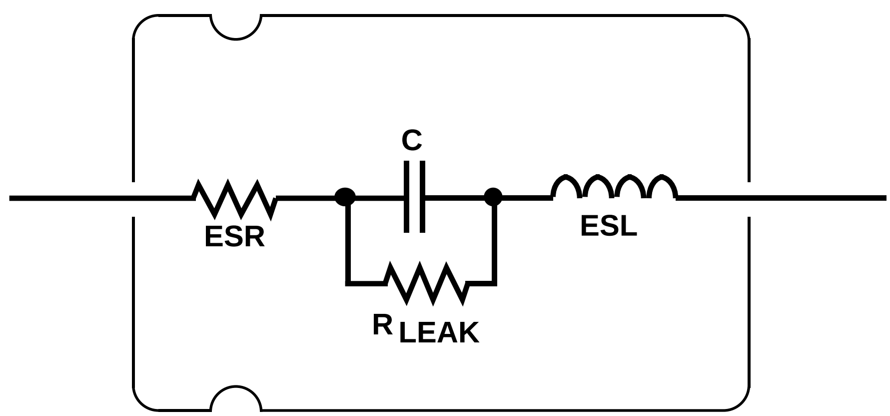


FIGURE 13. MODEL OF A REAL CAPACITOR

The most important elements are the **Equivalent Series Resistance (ESR)** and **Effective Series Inductance (ESL)**.

The **ESL limits a capacitors effectiveness at high frequencies**, and is the primary reason electrolytic capacitors must be bypassed by good RF capacitors in switching regulator applications (ceramic and film types are often used).

The **ESR is the primary cause of regulator loop instability in both linear LDO regulators and switching regulators**. In order to understand this, a brief review of loop theory will be presented to illustrate the effect of ESR on loop response.

REGULATOR LOOP RESPONSE

The loop response of a typical regulator is shown in Figure 14. The most important point to realize is that **for a stable loop, the gain must cross below 0 dB before the phase angle reaches 180**°.

A phase angle of 180° means that the signal being fed back around the loop is actually positive feedback, and will cause oscillations to occur.

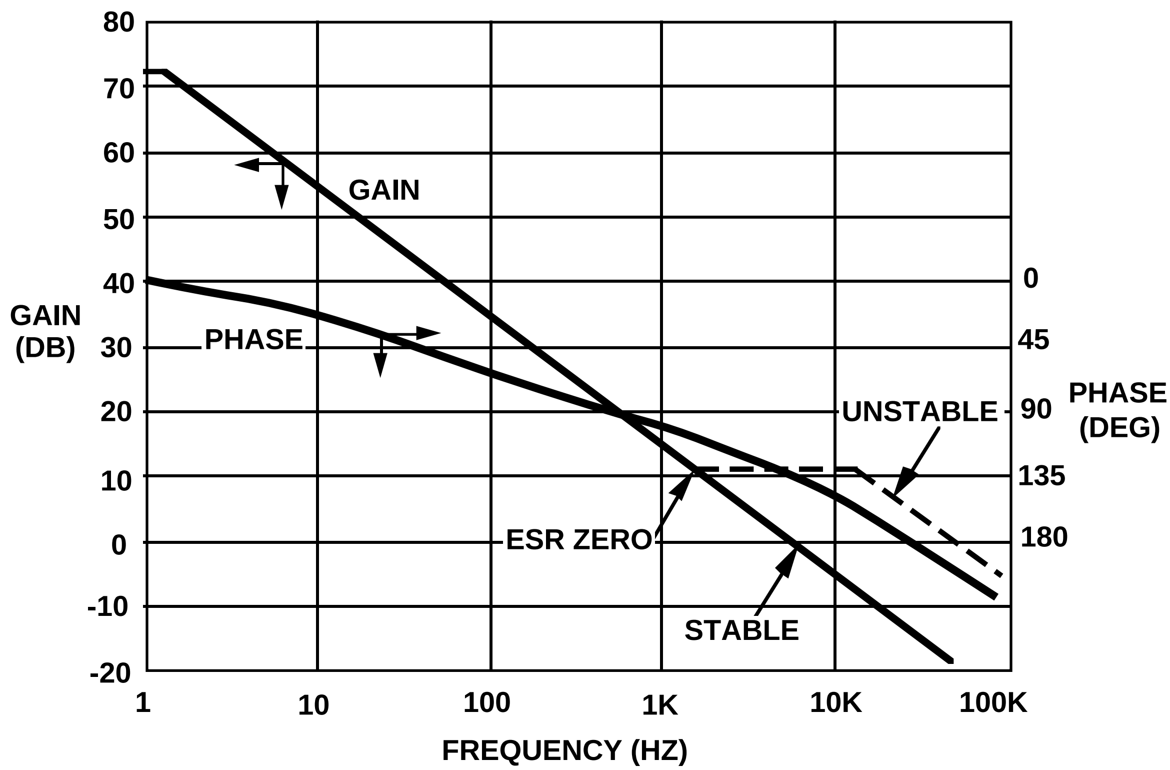


FIGURE 14. LOOP GAIN PLOT

(Note: In reality, a phase margin of 45° is usually required for good stability, which means it is advisable to get a 0 dB crossover before the phase angle reaches 135°).

In an LDO regulator, the output capacitor is required to force the gain to roll off fast enough to meet the stability requirements (a standard NPN regulator is internally compensated, and usually needs **no** output capacitor for stability).

As shown in Figure 14, the ESR of the output capacitor causes an unwanted "zero" in the response, which delays the 0 dB crossover point. **If the ESR is large enough, the "zero frequency" gets low enough to cause regulator instability**.

The stability requirements for a specific regulator will be listed on the data sheet for the part. In some cases, **a range is given which requires that the ESR be within the minimum and maximum limits**. In the newer parts, only a maximum limit must be met (which makes selecting a capacitor much easier).

TEMPERATURE DEPENDENCE OF ESR

Having now established the necessity of controlling the ESR of the output capacitor on an LDO regulator (to keep the regulator from oscillating), we need to point out one very important thing: **ESR is not constant with temperatur**e.

Figure 15 shows a plot of ESR versus temperature for a typical aluminum electrolytic capacitor. The most important point to observe is how fast the ESR increases at low temperatures.

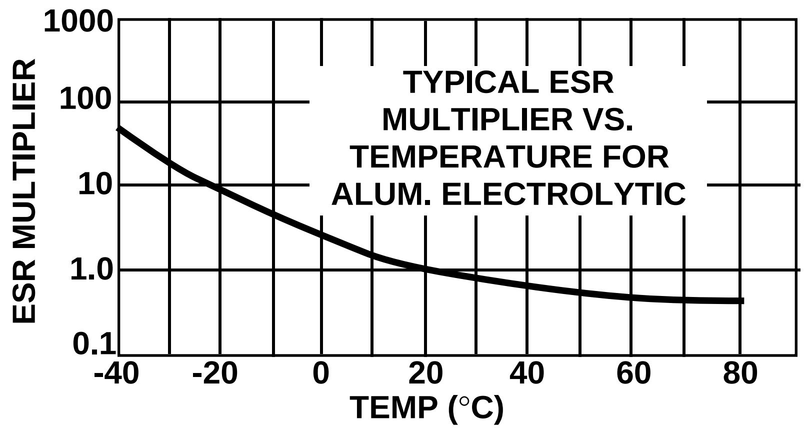


FIGURE 15. ALUMINUM ELECTROLYTIC ESR VS. TEMPERATURE

In cases where an LDO regulator must be operated below about -10 °C, it is sometimes not possible to find an aluminum electrolytic capacitor that can maintain an ESR within the acceptable range. Also, it is essential that the capacitor is specified to operate over the full temperature range: **some aluminum electrolytics are not usable below -20°C** (because their electrolyte freezes).

If the regulator has only a **maximum** limit which the ESR must not exceed, the aluminum electrolytic capacitor can be paralleled with a solid tantalum capacitor (which has a much lower ESR).

When two capacitors are in parallel, the effective ESR is the parallel of the two ESR values, which means the tantalum will help suppress the low-temperature ramp up seen in Figure 15. As a good rule, the tantalum should be selected so that its capacitance is about 20% of the aluminum electrolytic.

If the regulator has **both a maximum and minimum limit** (the ESR must stay in a specified range), it may be necessary to use a low value carbon film resistor placed in series with a low ESR capacitor (tantalum, film, or ceramic will work).

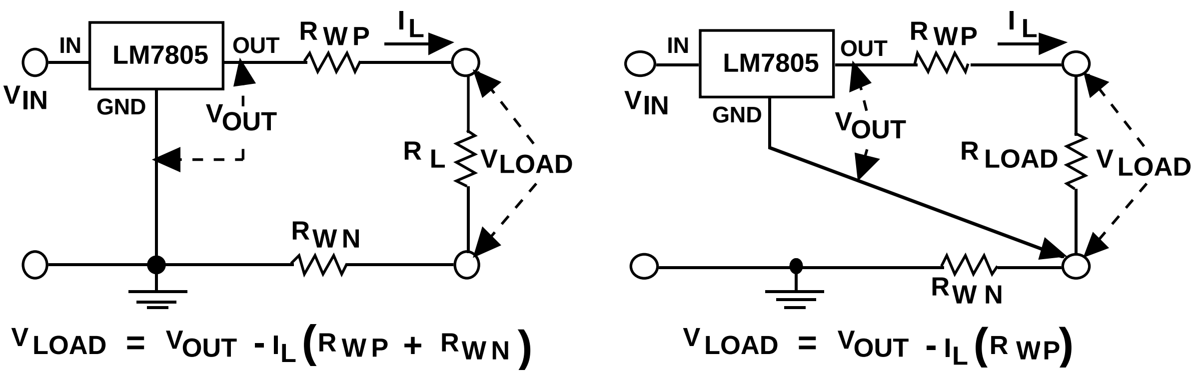
The best type of capacitor to use will depend upon how much total capacitance is required.

Load Regulation

The load regulation that a linear regulator **can** deliver is often much better than what is actually seen in the application due to voltage drops occurring along high-current paths. To understand how and why this occurs, we will look at examples of fixed and adjustable linear regulators.

FIXED OUTPUT REGULATORS

A typical application will be examined using an LM7805 three-terminal regulator (see Figure 16).



TYPICAL APPLICATION

IMPROVED LOAD REGULATION BY RETURNING GROUND PIN TO LOAD

FIGURE 16. LOAD REGULATION EFFECTS DUE TO WIRE DROPS

The user is most interested in the voltage at the load, but the LM7805 is regulating the voltage that appears between its output and ground pins. Any voltage drops that occur between the regulator pins and the load terminals reduce the voltage across the load (and degrade the load regulation).

In the typical application, V_{LOAD} is always less than V_{OUT} by the sum of the voltage drops appearing along the positive PC board trace (or wire) and the negative trace (or wire). The voltage drops along the leads are equal to the resistances (shown as R_{WP} and R_{WN}) multiplied times the load current.

This shows very clearly how trace resistance can cause "voltage errors" to occur at the load terminals, with the amount of "error" being directly related to the load current. In such cases, the regulation seen at the load would be considerably worse than the specification for the IC regulator.

This can be improved in two ways:

1) Move the regulator ground lead over and tie it directly to the negative load terminal, so that no other current can flow in this lead and cause voltage drops.

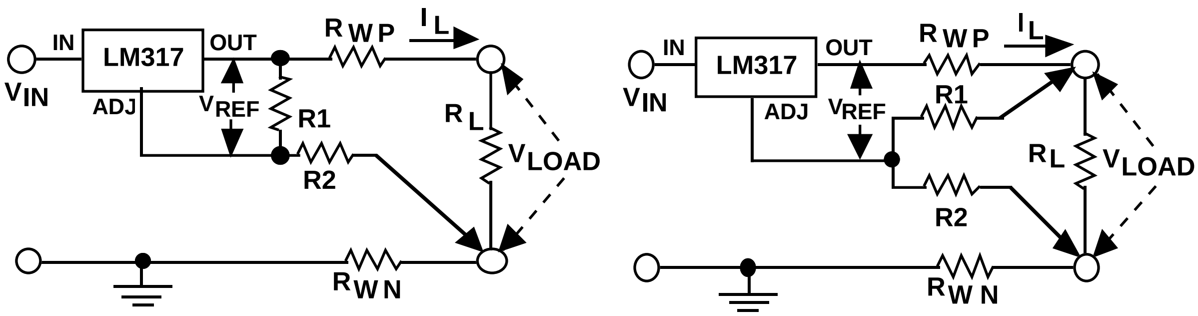
2) Minimize the drop in the positive lead by using the maximum possible conductor thickness, and place the IC regulator as near the load as is physically possible.

ADJUSTABLE OUTPUT REGULATORS

Adjustable linear regulators are different from fixed output types because an external resistive divider (along with the internal reference) is used to set the output voltage.

<u>Three-Terminal Regulators</u>

In the three-terminal adjustable regulators (like the **LM317**), **the reference voltage appears between the output pin and the adjust pin** (see Figure 17).



$$V_{LOAD} = V_{REF}\frac{(R1+R2)}{R1} - I_L(R_{WP})$$

$$V_{LOAD} = V_{REF}\frac{(R1+R2)}{R1} - I_L(R_{WP})\frac{(R1+R2)}{R1}$$

BEST LOAD REGULATION **NOT RECOMMENDED**

FIGURE 17. LOAD REGULATION EFFECTS USING LM317 REGULATOR

In the circuit for best load regulation, it is shown that the voltage appearing across the load is reduced from the nominal (no load) output voltage by **the voltage drop that results from the positive side trace resistance multiplied times the load current**.

As before, the **best performance** is obtained with the negative (ground) side of the resistive divider tied directly to the negative load terminal. This technique eliminates the drop in the negative high-current output trace (R_{WN}) from causing an additional decrease in V_{LOAD}.

It seems intuitively correct that an additional improvement would be obtained by tying the top side of the divider string to the positive load terminal, but this assumption is ABSOLUTELY WRONG.

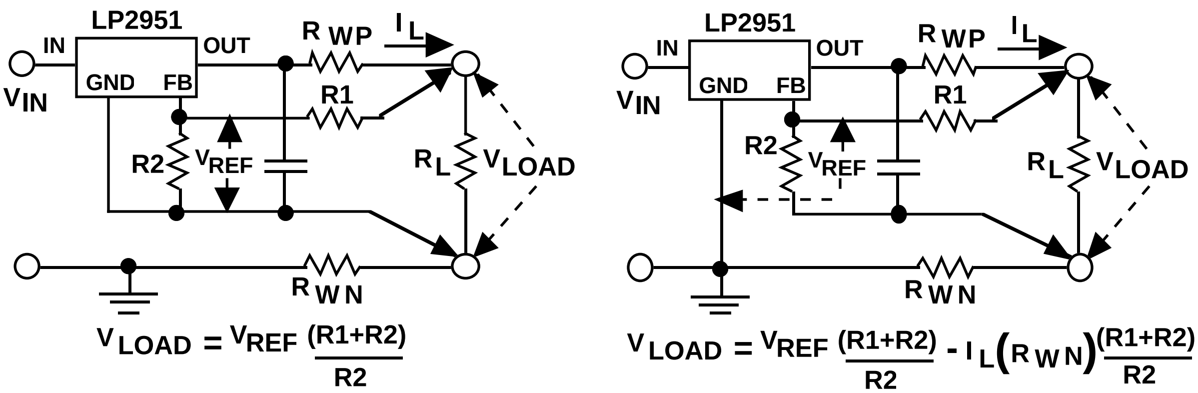
The voltage V_{REF} is used to force (set) a constant current through both R1 and R2, and the precision of the output voltage is directly related to the accuracy of this current. If R1 is tied to the positive load terminal, the voltage drop across R_{WP} is subtracted from V_{REF}, reducing the current through the divider.

The overall effect of the current change is that the voltage "error" is multiplied by the ratio of (1 + R2/R1), making the load regulation much worse.

Multi-pin Regulators

Adjustable regulators which are not limited to three pins have the advantage of using a ground pin, which allows the elimination of the output voltage error due to voltage drops along the output traces.

An example of such a regulator is the LP2951, a multi-function 250mA LDO regulator that can be adjusted to output voltages from 1.23V to 29V. In Figure 18, we see an LP2951 in a typical application. The voltage error at the load due to trace voltage drops is eliminated in the left-hand figure.



BEST LOAD REGULATION **NOT RECOMMENDED**

FIGURE 18. ELIMINATING LOAD REGULATION EFFECTS IN THE LP2951 REGULATOR

Note that the reference voltage in the LP2951 is regulated with respect to the ground pin, unlike the three-terminal adjustable regulators which have no ground pin. The discussion of this application is equally applicable to any regulator whose reference is regulated against ground.

In the left-hand figure, the trace voltage errors have been eliminated by tying the sense points of the resistive divider to the load terminals. **Important**: if this remote-sense method is used, the **ground pin must also be tied to the negative load terminal to prevent significant errors in V_{LOAD}** (see the right-hand figure).

If the ground pin and the lower sense point of R2 are separated, the voltage between these two points is multiplied by the ratio of (1+R1/R2) and appears as an error in the voltage V_{LOAD}. Since this error voltage is load current dependent, the voltage V_{LOAD} will also change with load current, resulting in poor load regulation.

For best load regulation, R2 should be located near the regulator with the ground pin tied directly to it. Then a single trace should be run to the negative load terminal, remembering that the trace size should be sufficient to assure a negligible voltage drop will occur along this lead when the part is conducting its maximum ground pin current (ground pin current can be as high as 45 mA in a 1A LDO regulator).

The Carrot in LDO Regulator Ground Pin Current

Many (but not all) LDO regulators have a characteristic in their ground pin current referred to as the "carrot". The carrot is a point in the ground pin current that spikes up as the input voltage is reduced (see Figure 19).

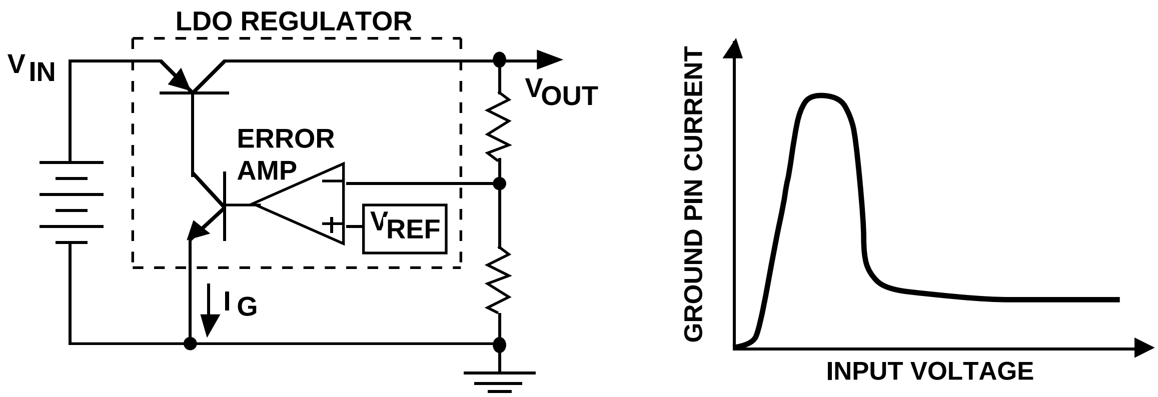


FIGURE 19. LDO REGULATOR WITH "CARROT"

The error amplifier in a regulator always tries to force the output to be the right voltage by adjusting the current through the pass device (in this case, the PNP transistor).

As the input voltage is reduced (and the voltage across the pass transistor decreases) the current gain of the PNP begins to drop. To maintain the correct output voltage, the error amplifier has to drive the base of the PNP harder to supply the same load current. The PNP base drive current leaves the regulator as ground pin current.

As the input voltage drops further, the regulator will approach dropout, causing the error amplifier to drive the PNP base with maximum current (this is the top of the carrot). This value of current may be 3 or 4 times the maximum ground pin current that is required to drive full rated load current with 5V across the pass transistor.

The carrot is recognized as an undesirable characteristic, since the additional ground pin current must be supplied by the source, but does not power the load (it just heats up the regulator).

In the newer LDO regulators, circuitry was built in to prevent this ground pin spike from occurring. For example, the **LP2951** (and all of the products in that family) have only a negligible increase in ground pin current as the input voltage crosses through the range where dropout is occurring.

SWITCHING REGULATORS

Introduction

The switching regulator is increasing in popularity because it offers the advantages of higher power conversion efficiency and increased design flexibility (multiple output voltages of different polarities can be generated from a single input voltage).

This paper will detail the operating principles of the four most commonly used switching converter types:

Buck: used the reduce a DC voltage to a lower DC voltage.

Boost: provides an output voltage that is higher than the input.

Buck-Boost (invert): an output voltage is generated opposite in polarity to the input.

Flyback: an output voltage that is less than or greater than the input can be generated, as well as multiple outputs.

Also, some multiple-transistor converter topologies will be presented:

Push-Pull: A two-transistor converter that is especially efficient at low input voltages.

Half-Bridge: A two-transistor converter used in many off-line applications.

Full-Bridge: A four-transistor converter (usually used in off-line designs) that can generate the highest output power of all the types listed.

Application information will be provided along with circuit examples that illustrate some applications of Buck, Boost, and Flyback regulators.

Switching Fundamentals

Before beginning explanations of converter theory, some basic elements of power conversion will be presented:

THE LAW OF INDUCTANCE

If a voltage is forced across an inductor, a current will flow through that inductor (and this current will vary with time). Note that the **current flowing in an inductor will be time-varying even if the forcing voltage is constant**.

It is equally correct to say that if a time-varying current is forced to flow in an inductor, a voltage across the inductor will result.

The fundamental law that defines the relationship between the voltage and current in an inductor is given by the equation:

$$v = L \ (di/dt)$$

Two important characteristics of an inductor that follow directly from the law of inductance are:

1) A voltage across an inductor results **only** from **a current that changes with time**. A **steady (DC) current flowing in an inductor causes no voltage across it** (except for the tiny voltage drop across the copper used in the windings).

2) A **current flowing in an inductor can not change value instantly** (in zero time), as this would require infinite voltage to force it to happen. However, **the faster the current is changed in an inductor, the larger the resulting voltage will be**.

Note: Unlike the current flowing in the inductor, the **voltage across it <u>can</u> change instantly** (in zero time).

The principles of inductance are illustrated by the information contained in Figure 25.

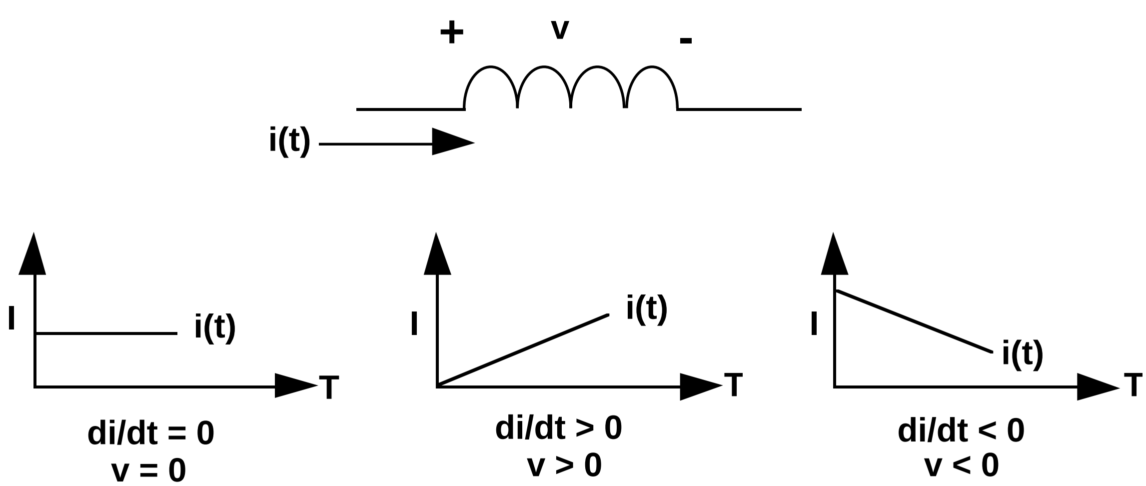


FIGURE 25. INDUCTOR VOLTAGE/CURRENT RELATIONSHIP

The important parameter is the di/dt term, which is simply a measure of how the current changes with time. When the current is plotted versus time, the **value of di/dt is defined as the slope of the current plot at any given point**.

The graph on the left shows that current which is constant with time has a di/dt value of zero, and results in no voltage across the inductor.

The center graph shows that a current which is increasing with time has a positive di/dt value, resulting in a positive inductor voltage.

Current that decreases with time (shown in the right-hand graph) gives a negative value for di/dt and inductor voltage.

It is important to note that a **linear current ramp in an inductor** (either up or down) **occurs only when it has a constant voltage across it**.

TRANSFORMER OPERATION

A transformer is a device that has two or more magnetically-coupled windings. The basic operation is shown in Figure 26.

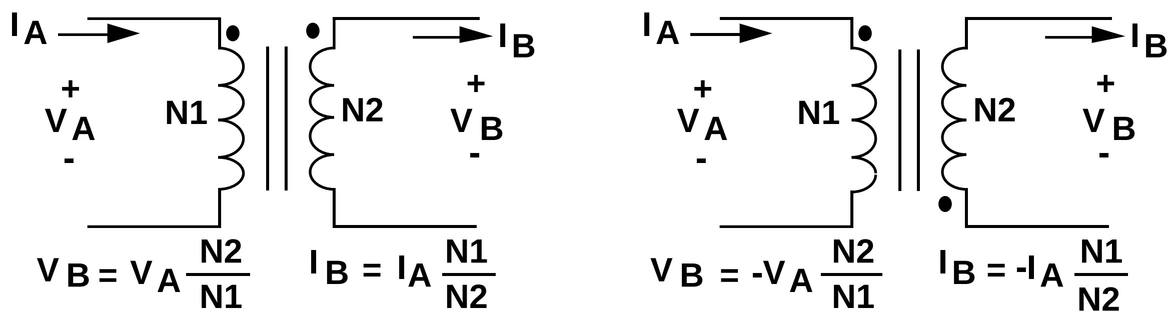


FIGURE 26. TRANSFORMER THEORY

The action of a transformer is such that a time-varying (AC) voltage or current is transformed to a higher or lower value, as set by the transformer turns ratio. The transformer does not add power, so it follows that the power (V X I) on either side must be constant. That is the reason that the **winding with more turns has higher voltage but lower current**, while the **winding with less turns has lower voltage but higher current**.

The **dot on a transformer winding identifies its polarity** with respect to another winding, and **reversing the dot results in inverting the polarity**.

Example of Transformer Operation:

An excellent example of how a transformer works can be found under the hood of your car, where a transformer is used to generate the 40 kV that fires your cars spark plugs (see Figure 27).

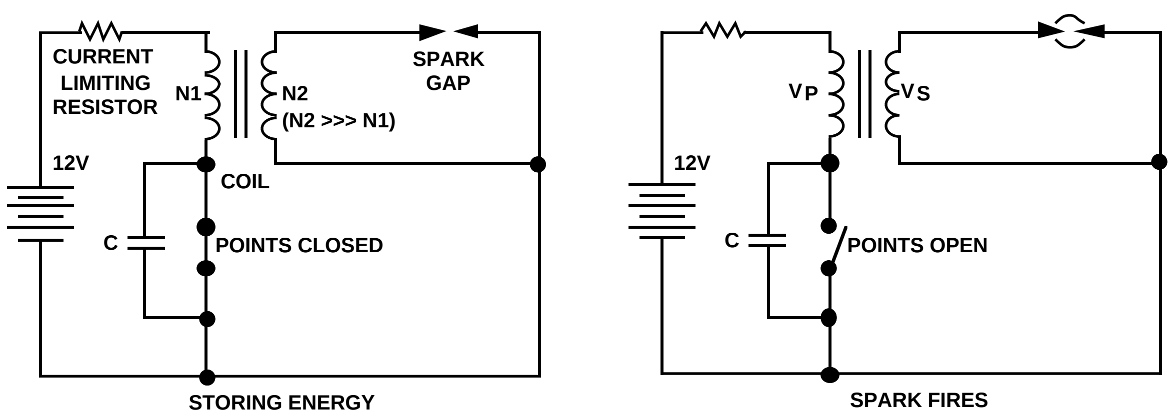


FIGURE 27. SPARK FIRING CIRCUIT

The "coil" used to generate the spark voltage is actually a transformer, with a very high secondary-to-primary turns ratio.

When the points first close, current starts to flow in the primary winding and eventually reaches the final value set by the 12V battery and the current limiting resistor. At this time, the current flow is a fixed DC value, which means no voltage is generated across either winding of the transformer.

When the points open, the current in the primary winding collapses very quickly, causing a large voltage to appear across this winding. This voltage on the primary is magnetically coupled to (and stepped up by) the secondary winding, generating a voltage of 30 kV - 40 kV on the secondary side.

As explained previously, the law of inductance says that it is not possible to instantly break the current flowing in an inductor (because an infinite voltage would be required to make it happen).

This principle is what causes the arcing across the contacts used in switches that are in circuits with highly inductive loads. When the switch just begins to open, the high voltage generated allows electrons to jump the air gap so that the current flow does not actually stop instantly. Placing a capacitor across the contacts helps to reduce this arcing effect.

In the automobile ignition, a capacitor is placed across the points to minimize damage due to arcing when the points "break" the current flowing in the low-voltage coil winding (in car manuals, this capacitor is referred to as a "condenser").

PULSE WIDTH MODULATION (PWM)

All of the switching converters that will be covered in this paper use a form of output voltage regulation known as **Pulse Width Modulation (PWM)**. Put simply, the feedback loop adjusts (corrects) the output voltage by changing the ON time of the switching element in the converter.

As an example of how PWM works, we will examine the result of applying a series of square wave pulses to an L-C filter (see Figure 28).

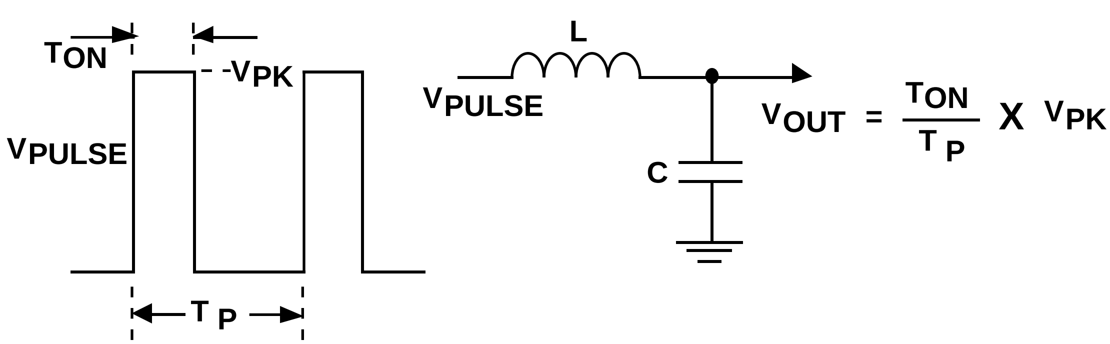


FIGURE 28. BASIC PRINCIPLES OF PWM

The series of square wave pulses is filtered and provides a **DC output voltage that is equal to the peak pulse amplitude multiplied times the duty cycle** (duty cycle is defined as the switch ON time divided by the total period).

This relationship explains how the output voltage can be directly controlled by changing the ON time of the switch.

Switching Converter Topologies

The most commonly used DC-DC converter circuits will now be presented along with the basic principles of operation.

BUCK REGULATOR

The most commonly used switching converter is the Buck, which is used to down-convert a DC voltage to a lower DC voltage of the same polarity. This is essential in systems that use distributed power rails (like 24V to 48V), which must be locally converted to 15V, 12V or 5V with very little power loss.

The Buck converter uses a transistor as a switch that alternately connects and disconnects the input voltage to an inductor (see Figure 29).

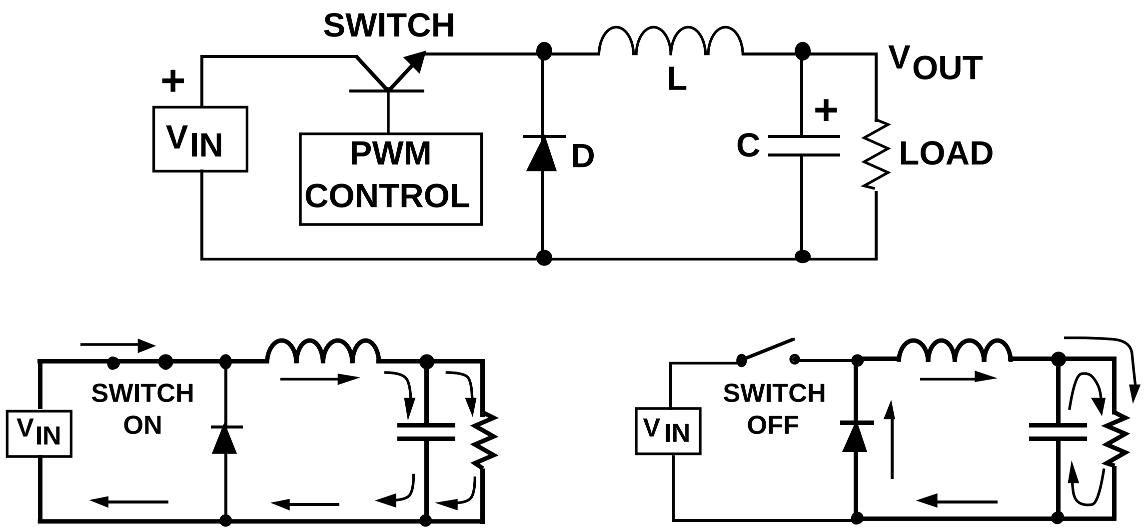


FIGURE 29. BUCK REGULATOR

The lower diagrams show the current flow paths (shown as the heavy lines) when the switch is on and off.

When the switch turns on, the input voltage is connected to the inductor. The difference between the input and output voltages is then forced across the inductor, causing current through the inductor to increase.

During the **on time**, the inductor current flows into both the load and the output capacitor (the capacitor **charges** during this time).

When the switch is turned off, the input voltage applied to the inductor is removed. However, since the current in an inductor can not change instantly, the voltage across the inductor will adjust to hold the current constant.

The input end of the inductor is forced negative in voltage by the decreasing current, eventually reaching the point where the diode is turned on. The inductor current then flows through the load and back through the diode.

The capacitor discharges into the load during the off time, contributing to the total current being supplied to the load (the total load current during the switch off time is the sum of the inductor and capacitor current).

The shape of the current flowing in the inductor is similar to Figure 30.

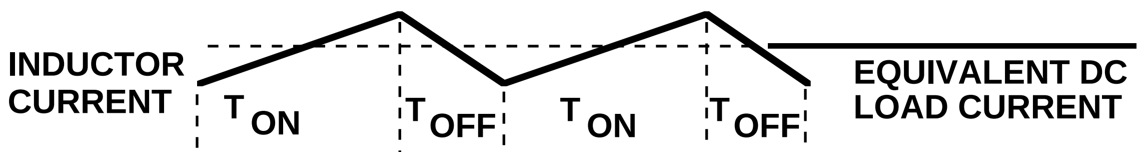


FIGURE 30. BUCK REGULATOR INDUCTOR CURRENT

As explained, the current through the inductor ramps up when the switch is on, and ramps down when the switch is off. The DC load current from the regulated output is the average value of the inductor current.

The peak-to-peak difference in the inductor current waveform is referred to as the **inductor ripple current**, and the inductor is typically selected large enough to keep this ripple current less than 20% to 30% of the rated DC current.

CONTINUOUS vs. DISCONTINUOUS OPERATION

In most Buck regulator applications, the inductor current never drops to zero during full-load operation (this is defined as **continuous mode operation**). Overall performance is usually better using continuous mode, and it allows maximum output power to be obtained from a given input voltage and switch current rating.

In applications where the maximum load current is fairly low, it can be advantageous to design for discontinuous mode operation. In these cases, operating in discontinuous mode can result in a smaller overall converter size (because a smaller inductor can be used).

Discontinuous mode operation at lower load current values is generally harmless, and **even converters designed for continuous mode operation at full load will become discontinuous as the load current is decreased** (usually causing no problems).

BOOST REGULATOR

The Boost regulator takes a DC input voltage and produces a DC output voltage that is higher in value than the input (but of the same polarity). The Boost regulator is shown in Figure 31, along with details showing the path of current flow during the switch on and off time.

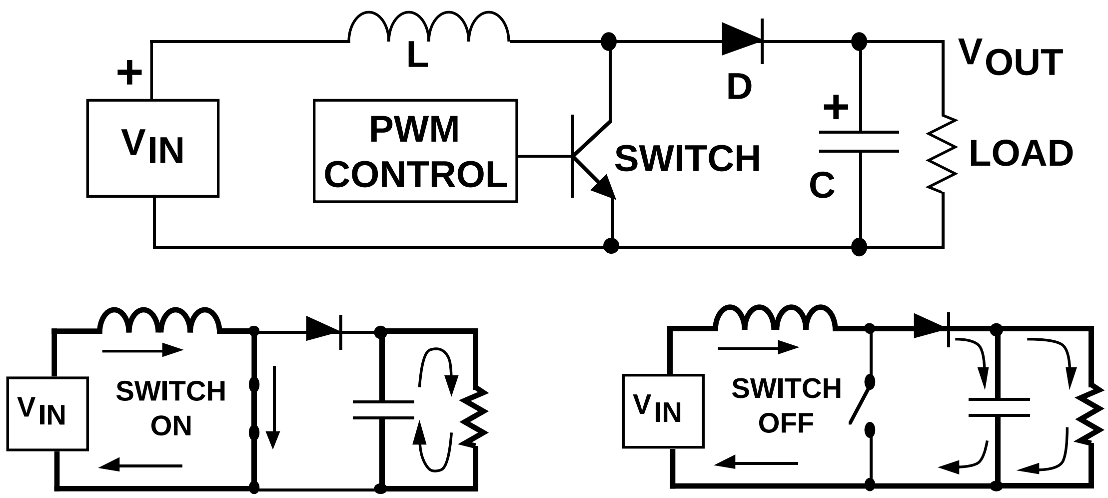


FIGURE 31. BOOST REGULATOR

Whenever the switch is on, the input voltage is forced across the inductor which causes the current through it to increase (ramp up).

When the switch is off, the decreasing inductor current forces the "switch" end of the inductor to swing positive. This forward biases the diode, allowing the capacitor to charge up to a voltage that is higher than the input voltage.

During steady-state operation, the inductor current flows into both the output capacitor and the load during the switch off time. When the switch is on, the load current is supplied only by the capacitor.

OUTPUT CURRENT AND LOAD POWER

An important design consideration in the Boost regulator is that the **output load current and the switch current are not equal**, and the **maximum available load current is always less than the current rating of the switch transistor**.

It should be noted that the **maximum total power available for conversion in <u>any</u> regulator** is equal to the **input voltage multiplied times the maximum average input current** (which is **less than** the current rating of the switch transistor).

Since the output voltage of the Boost is **higher than the input voltage**, it follows that the **output current must be lower than the input current**.

BUCK-BOOST (INVERTING) REGULATOR

The Buck-Boost or Inverting regulator takes a DC input voltage and produces a DC output voltage that is opposite in polarity to the input. The negative output voltage can be either larger or smaller in magnitude than the input voltage.

The Inverting regulator is shown in Figure 32.

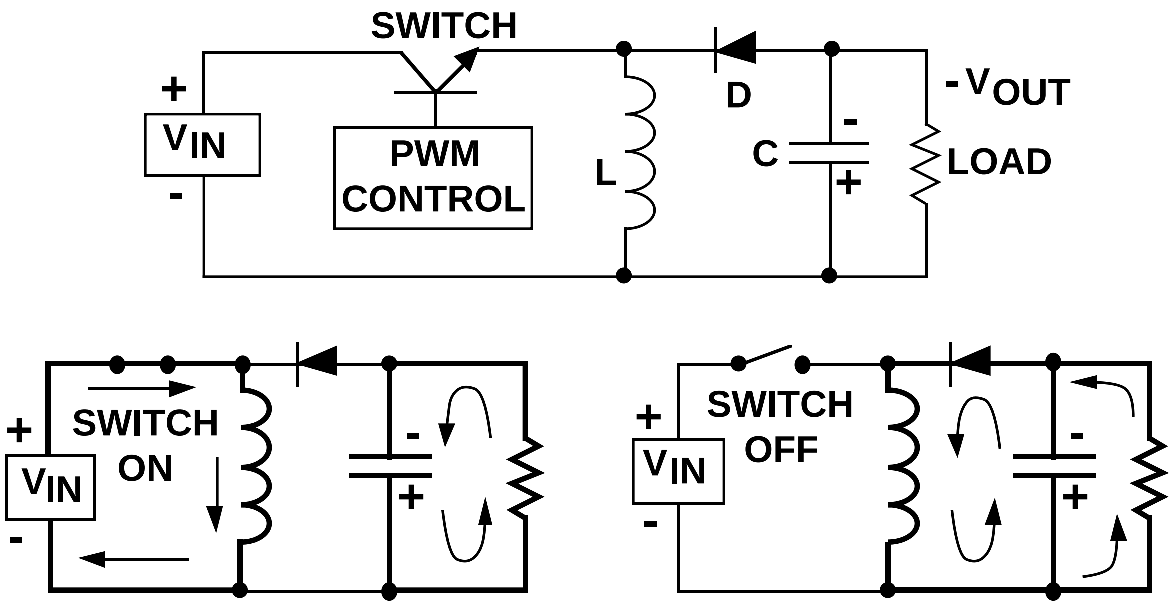


FIGURE 32. BUCK-BOOST (INVERTING) REGULATOR

When the switch is on, the input voltage is forced across the inductor, causing an increasing current flow through it. During the on time, the discharge of the output capacitor is the only source of load current.

This requires that the charge lost from the output capacitor during the on time be replenished during the off time.

When the switch turns off, the decreasing current flow in the inductor causes the voltage at the diode end to swing negative. This action turns on the diode, allowing the current in the inductor to supply **both the output capacitor and the load**.

As shown, **the load current is supplied by inductor when the switch is off, and by the output capacitor when the switch is on**.

FLYBACK REGULATOR

The Flyback is the most versatile of all the topologies, allowing the designer to create one or more output voltages, some of which may be opposite in polarity.

Flyback converters have gained popularity in battery-powered systems, where a single voltage must be converted into the required system voltages (for example, +5V, +12V and -12V) with very high power conversion efficiency.

The basic single-output flyback converter is shown in Figure 33.

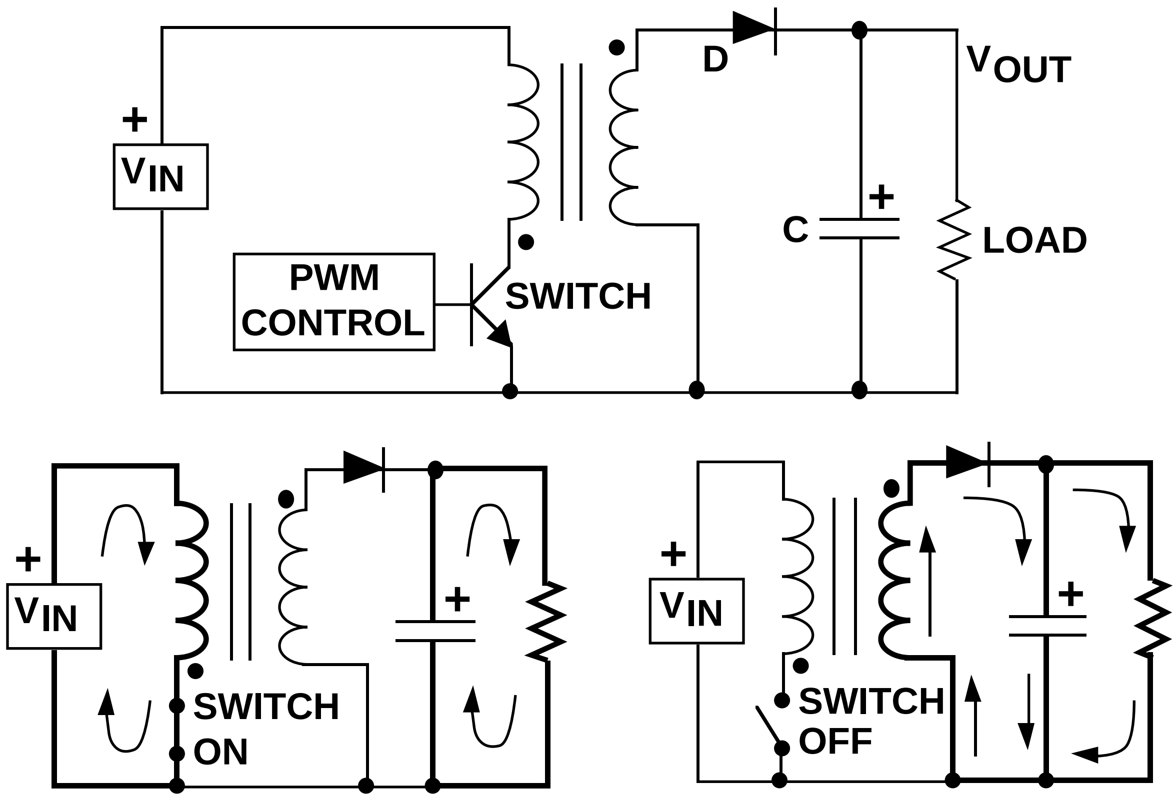


FIGURE 33. SINGLE-OUTPUT FLYBACK REGULATOR

The most important feature of the Flyback regulator is the transformer phasing, as shown by the dots on the primary and secondary windings.

When the switch is on, the input voltage is forced across the transformer primary which causes an increasing flow of current through it.

Note that the **polarity of the voltage on the primary is dot-negative** (more negative at the dotted end), causing a voltage with the same polarity to appear at the transformer secondary (the magnitude of the secondary voltage is set by the transformer seconday-to-primary turns ratio).

The dot-negative voltage appearing across the secondary winding turns off the diode, preventing current flow in the secondary winding during the switch on time. During this time, the load current must be supplied by the output capacitor alone.

When the switch turns off, the decreasing current flow in the primary causes the voltage at the dot end to swing positive. At the same time, the primary voltage is reflected to the secondary with the same polarity. The dot-positive voltage occurring across the secondary winding turns on the diode, allowing current to flow into both the load and the output capacitor. The output capacitor charge lost to the load during the switch on time is replenished during the switch off time.

Flyback converters operate in either **continuous mode** (where the secondary current is always >0) or **discontinuous mode** (where the secondary current falls to zero on each cycle).

GENERATING MULTIPLE OUTPUTS

Another big advantage of a Flyback is the capability of providing multiple outputs (see Figure 34). In such applications, one of the outputs (usually the highest current) is selected to provide PWM feedback to the control loop, **which means this output is directly regulated.**

The other secondary winding(s) are **indirectly regulated**, as their pulse widths will follow the regulated winding. The load regulation on the unregulated secondaries is not great (typically 5 - 10%), but is adequate for many applications.

If tighter regulation is needed on the lower current secondaries, an LDO post-regulator is an excellent solution. The secondary voltage is set about 1V above the desired output voltage, and the LDO provides excellent output regulation with very little loss of efficiency.

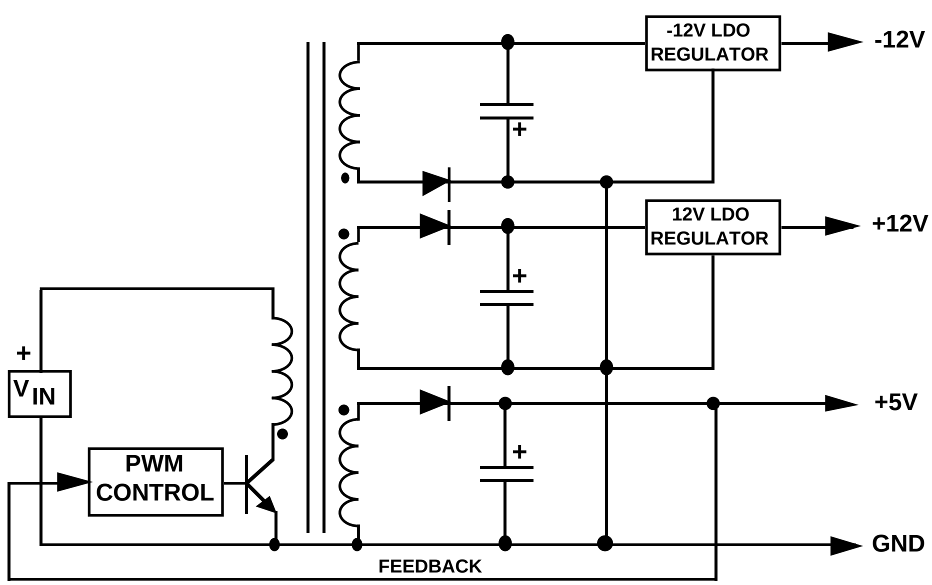


FIGURE 34. TYPICAL MULTIPLE-OUTPUT FLYBACK

PUSH-PULL CONVERTER

The Push-Pull converter uses two to transistors perform DC-DC conversion (see Figure 35).

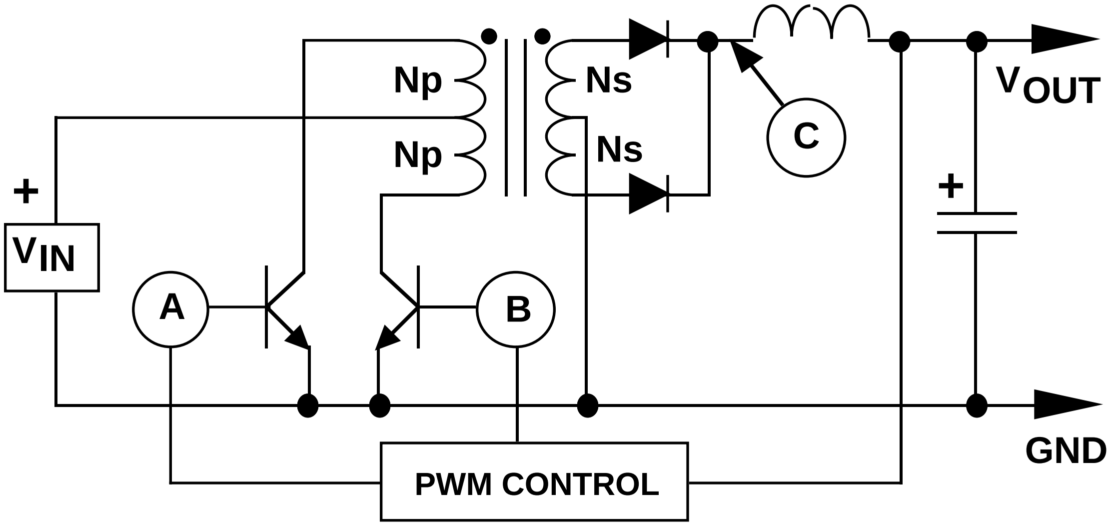


FIGURE 35. PUSH-PULL CONVERTER

The converter operates by turning on each transistor on alternate cycles (the two transistors are **never** on at the same time). **Transformer secondary current flows at the same time as primary current** (when either of the switches is on).

For example, when transistor "A" is turned on, the input voltage is forced across the upper primary winding with dot-negative polarity. On the secondary side, a dot-negative voltage will appear across the winding which turns on the bottom diode. This allows current to flow into the inductor to supply both the output capacitor and the load.

When transistor "B" is on, the input voltage is forced across the lower primary winding with dot-positive polarity. The same voltage polarity on the secondary turns on the top diode, and current flows into the output capacitor and the load.

An important characteristic of a Push-Pull converter is that the switch transistors have to be able the stand off more than twice the input voltage: when one transistor is on (and the input voltage is forced across one primary winding) the same magnitude voltage is induced across the other primary winding, but it is "floating" on top of the input voltage. This puts the collector of the turned-off transistor at **twice** the input voltage with respect to ground.

The "double input voltage" rating requirement of the switch transistors means the Push-Pull converter is best suited for lower input voltage applications. It has been widely used in converters operating in 12V and 24V battery-powered systems.

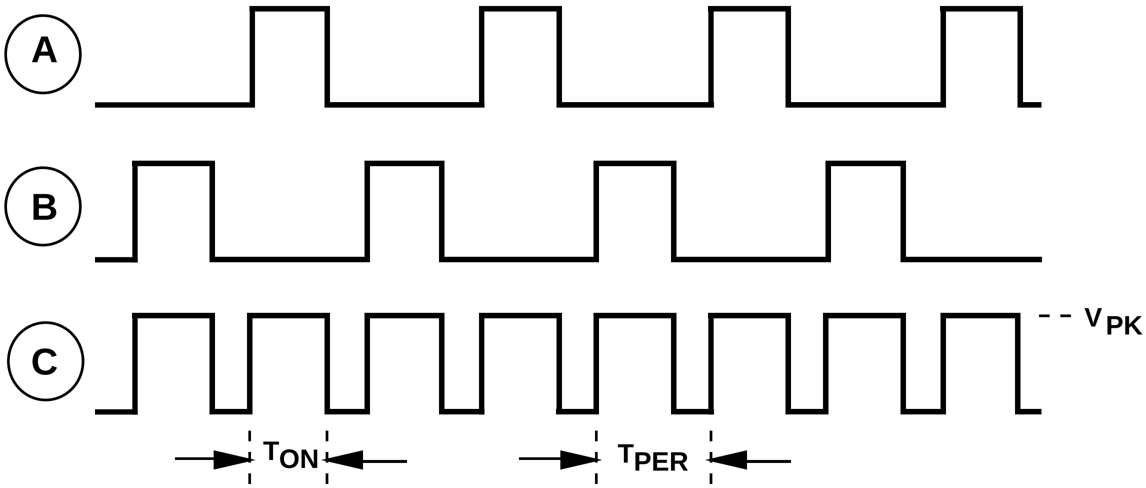


FIGURE 36. TIMING DIAGRAM FOR PUSH-PULL CONVERTER

Figure 36 shows a timing diagram which details the relationship of the input and output pulses.

It is important to note that frequency of the secondary side voltage pulses is twice the frequency of operation of the PWM controller driving the two transistors. For example, if the PWM control chip was set up to operate at 50 kHz on the primary side, the frequency of the secondary pulses would be 100 kHz.

The DC output voltage is given by the equation:

$$V_{OUT} = V_{PK} \times (T_{ON} / T_{PER})$$

The peak amplitude of the secondary pulses (V_{PK}) is given by:

$$V_{PK} = (V_{IN} - V_{SWITCH}) \times (N_S / N_P) - V_{RECT}$$

This highlights why the Push-Pull converter is well-suited for low voltage converters. The voltage forced across each primary winding (which provides the power for conversion) is the full input voltage minus only the saturation voltage of the switch.

If MOS-FET power switches are used, the voltage drop across the switches can be made extremely small, resulting in very high utilization of the available input voltage.

Another advantage of the Push-Pull converter is that it can also generate multiple output voltages (by adding more secondary windings), some of which may be negative in polarity. This allows a power supply operated from a single battery to provide all of the voltages necessary for system operation.

A disadvantage of Push-Pull converters is that they require very good matching of the switch transistors to prevent unequal on times, since this will result in saturation of the transformer core (and failure of the converter).

HALF-BRIDGE CONVERTER

The Half-Bridge is a two-transistor converter frequently used in high-power designs. It is well-suited for applications requiring load power in the range of 500W to 1500W, and is almost always operated directly from the AC line.

Off-line operation means that no large 60 Hz power transformer is used, eliminating the heaviest and costliest component of a typical transformer-powered supply. All of the transformers in the Half-Bridge used for power conversion operate at the switching frequency (typically 50 kHz or higher) which means they can be very small and efficient.

A very important advantage of the Half-Bridge is **input-to-output isolation** (the regulated DC output is electrically isolated from the AC line). But, this means that all of the PWM control circuitry must be referenced to the DC output ground.

The voltage to run the control circuits is usually generated from a DC rail that is powered by a small 60 Hz transformer feeding a three-terminal regulator. In some designs requiring extremely high efficiency, the switcher output takes over and provides internal power after the start-up period.

The switch transistor drive circuitry must be isolated from the transistors, requiring the use of base drive transformers. The added complexity of the base drive circuitry is a disadvantage of using the Half-Bridge design.

If a 230 VAC line voltage is rectified by a full-wave bridge and filtered by a capacitor, an unregulated DC voltage of about 300V will be available for DC-DC conversion. If 115 VAC is used, a voltage doubler circuit is typically used to generate the 300V rail.

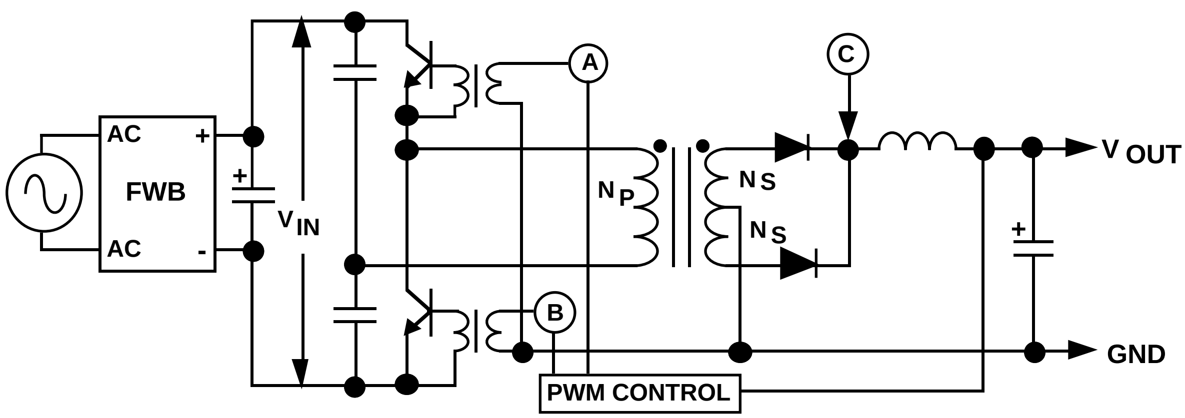


FIGURE 37. HALF-BRIDGE CONVERTER

The basic Half-bridge converter is shown in Figure 37. A capacitive divider is tied directly across the unregulated DC input voltage, providing a reference voltage of $1/2V_{IN}$ for one end of the transformer primary winding. The other end of the primary is actively driven up and down as the transistors alternately turn on and off.

The switch transistors force one-half of the input voltage across the primary winding during the switch on time, reversing polarity as the transistors alternate. The switching transistors are **never** on at the same time, or they would be destroyed (since they are tied directly across V_{IN}). The timing diagram for the Half-Bridge converter is shown in Figure 38 (it is the same as the Push-Pull).

When the "A" transistor is on, a dot-positive voltage is forced across the primary winding and reflected on the secondary side (with the magnitude being set by the transformer turns ratio). The dot-positive secondary voltage turns on the upper rectifier diode, supplying current to both the output capacitor and the load.

When the "A" transistor turns off and the "B" transistor turns on, the polarity of the primary voltage is reversed. The secondary voltage polarity is also reversed, turning on the lower diode (which supplies current to the output capacitor and the load).

In a Half-Bridge converter, **primary and secondary current flow in the transformer at the same time** (when either transistor is on), supplying the load current and charging the output capacitor. The output capacitor discharges into the load **only during the time when both transistors are off**.

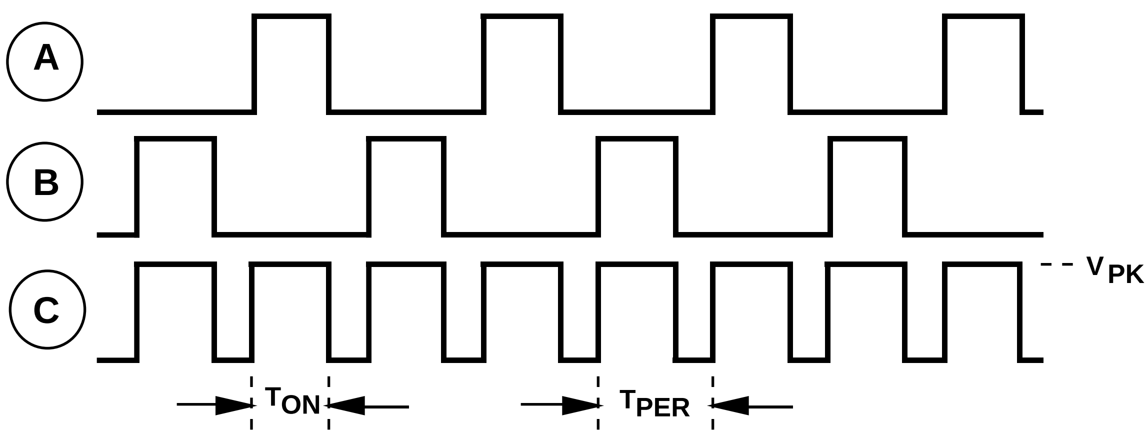


FIGURE 38. TIMING DIAGRAM FOR HALF-BRIDGE CONVERTER

It can be seen that the voltage pulses on the transformer secondary side (applied to the L-C filter) are occurring at twice the frequency of the PWM converter which supplies the drive pulses for the switching transistors.

The output voltage is again given by:

$$V_{OUT} = V_{PK} \times (T_{ON} / T_{PER})$$

The peak amplitude of the secondary pulses (V_{PK}) is given by:

$$V_{PK} = (1/2\ V_{IN} - V_{SWITCH}) \times (N_S / N_P) - V_{RECT}$$

FULL-BRIDGE CONVERTER

The Full-Bridge converter requires a total of four switching transistors to perform DC-DC conversion. The full bridge is most often seen in applications that are powered directly from the AC line, providing load power of 1 kW to 3 kW.

Operating off-line, the Full Bridge converter typically uses about 300V of unregulated DC voltage for power conversion (the voltage that is obtained when a standard 230 VAC line is rectified and filtered).

An important feature of this design is the isolation from the AC line provided by the switching transformer. The PWM control circuitry is referenced to the the output ground, requiring a dedicated voltage rail (usually powered from a small 60 Hz transformer) to run the control circuits.

The base drive voltages for the switch transistors (which are provided by the PWM chip) have to be transformer-coupled because of the required isolation.

Figure 39 shows a simplified schematic diagram of a Full-Bridge converter.

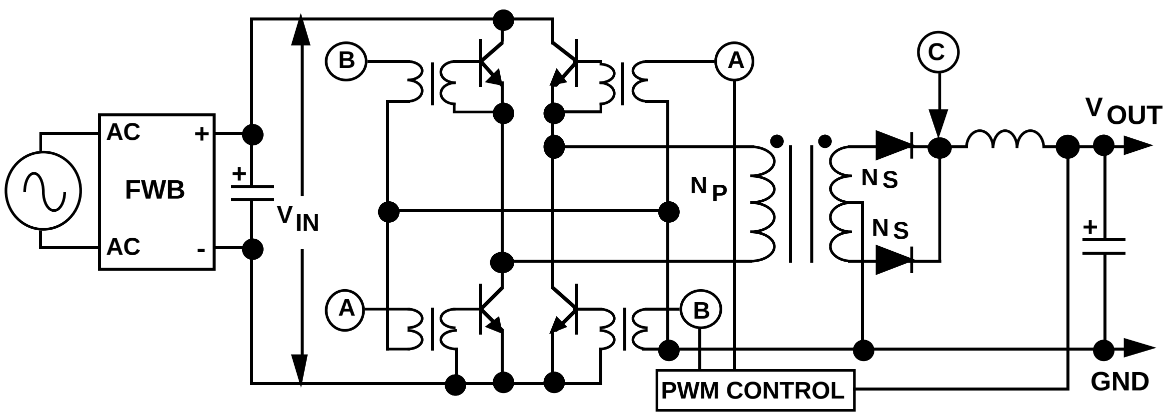


FIGURE 39. FULL BRIDGE CONVERTER

The transformer primary is driven by the **full voltage V_{IN}** when either of the transistor sets ("A" set or "B" set) turns on. The full input voltage utilization means the **Full-Bridge can produce the most load power of all the converter types**. The timing diagram is identical to the Half-Bridge, as shown in Figure 38.

Primary and secondary current flows in the transformer during the switch on times, while the output capacitor discharges into the load when both transistors are off.

The equation for the output voltage is (see Figure 38):

$$V_{OUT} = V_{PK} \text{ X } (T_{ON} / T_{PER})$$

The peak voltage of the transformer secondary pulses (V_{PK}) is given by:

$$V_{PK} = (V_{IN} - 2V_{SWITCH}) \text{ X } (N_S / N_P) - V_{RECT}$$

Application Hints For Switching Regulators

Application information will be provided on topics which will enhance the designers ability to maximize switching regulator performance.

Capacitor Parasitics Affecting Switching Regulator Performance

All capacitors contain parasitic elements which make their performance less than ideal (see Figure 40).

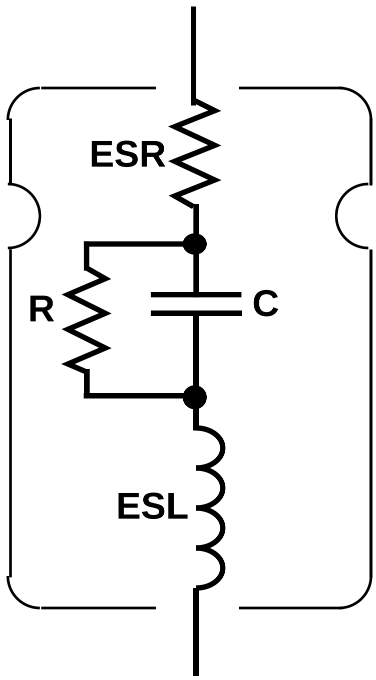


FIGURE 40.
CAPACITOR PARASITICS

Summary of Effects of Parasitics:

ESR: The ESR (**Equivalent Series Resistance**) causes internal heating due to power dissipation as the ripple current flows into and out of the capacitor. The capacitor can fail if ripple current exceeds maximum ratings.

Excessive output voltage ripple will result from high ESR**, and regulator loop instability is also possible**. ESR is highly dependent on temperature, increasing very quickly at temperatures below about 10 °C.

ESL: The ESL (**Effective Series Inductance**) limits the high frequency effectiveness of the capacitor. High ESL is the reason electrolytic capacitors need to be bypassed by film or ceramic capacitors to provide good high-frequency performance.

The ESR, ESL and C within the capacitor form a resonant circuit, whose frequency of resonance should be as high as possible. Switching regulators generate ripple voltages on their outputs with very high frequency (>10 MHz) components, which can cause ringing on the output voltage if the capacitor resonant frequency is low enough to be near these frequencies.

INPUT CAPACITORS

All of the switching converters in this paper (and the vast majority in use) operate as DC-DC converters that "chop" a DC input voltage at a very high frequency. As the converter switches, it has to draw current pulses from the input source. The **source impedance is extremely important**, as even a small amount of inductance can cause significant ringing and spiking on the voltage at the input of the converter.

The best practice is to **always provide adequate capacitive bypass as near as possible to the switching converter input**. For best results, an electrolytic is used with a film capacitor (and possibly a ceramic capacitor) in parallel for optimum high frequency bypassing.

OUTPUT CAPACITOR ESR EFFECTS

The primary function of the output capacitor in a switching regulator is filtering. As the converter operates, current must flow into and out of the output filter capacitor.

The ESR of the output capacitor directly affects the performance of the switching regulator. ESR is specified by the manufacturer on good quality capacitors, **but be certain that it is specified at the frequency of intended operation**.

General-purpose electrolytics usually only specify ESR at 120 Hz, but capacitors intended for high-frequency switching applications will have the ESR guaranteed at high frequency (like 20 kHz to 100 kHz).

Some ESR dependent parameters are:

<u>Ripple Voltage</u>: In most cases, **the majority of the output ripple voltage results from the ESR of the output capacitor**. If the ESR increases (as it will at low operating temperatures) the output ripple voltage will increase accordingly.

<u>Efficiency</u>: As the switching current flows into and out of the capacitor (through the ESR), power is dissipated internally. This "wasted" power reduces overall regulator efficiency, and **can also cause the capacitor to fail if the ripple current exceeds the maximum allowable specification for the capacitor**.

<u>Loop Stability</u>: The ESR of the output capacitor can affect regulator loop stability. Products such as the LM2575 and LM2577 are compensated for stability assuming the ESR of the output capacitor will stay within a specified range.

Keeping the ESR within the "stable" range is not always simple in designs that must operate over a wide temperature range. **The ESR of a typical aluminum electrolytic may increase by 40X as the temperature drops from 25°C to -40°C**.

In these cases, an aluminum electrolytic must be paralleled by another type of capacitor with a flatter ESR curve (like Tantalum or Film) so that the effective ESR (which is the parallel value of the two ESR's) stays within the allowable range.

Note: if operation **below -40°C** is necessary, aluminum electrolytics are probably not feasible for use.

BYPASS CAPACITORS

High-frequency bypass capacitors are always recommended on the supply pins of IC devices, but if the devices are used in assemblies near switching converters **bypass capacitors are absolutely required.**

The components which perform the high-speed switching (transistors and rectifiers) generate significant EMI that easily radiates into PC board traces and wire leads.

To assure proper circuit operation, all IC supply pins must be bypassed to a clean, low-inductance ground (for details on grounding, see next section).

Proper Grounding

The "ground" in a circuit is supposed to be at one potential, but in real life it is not. When ground currents flow through traces which have non-zero resistance, voltage differences will result at different points along the ground path.

In DC or low-frequency circuits, "ground management" is comparatively simple: the only parameter of critical importance is the **DC resistance of a conductor**, since that defines the voltage drop across it for a given current. In high-frequency circuits, it is the **inductance** of a trace or conductor that is much more important.

In switching converters, peak currents flow in high-frequency (> 50 kHz) pulses, which can cause severe problems if trace inductance is high. Much of the "ringing" and "spiking" seen on voltage waveforms in switching converters is the result of high current being switched through parasitic trace (or wire) inductance.

Current switching at high frequencies tends to flow near the surface of a conductor (this is called "skin effect"), which means that ground traces must be very wide on a PC board to avoid problems. It is usually best (when possible) to use one side of the PC board as a **ground plane**.

Figure 41 illustrates an example of a terrible layout:

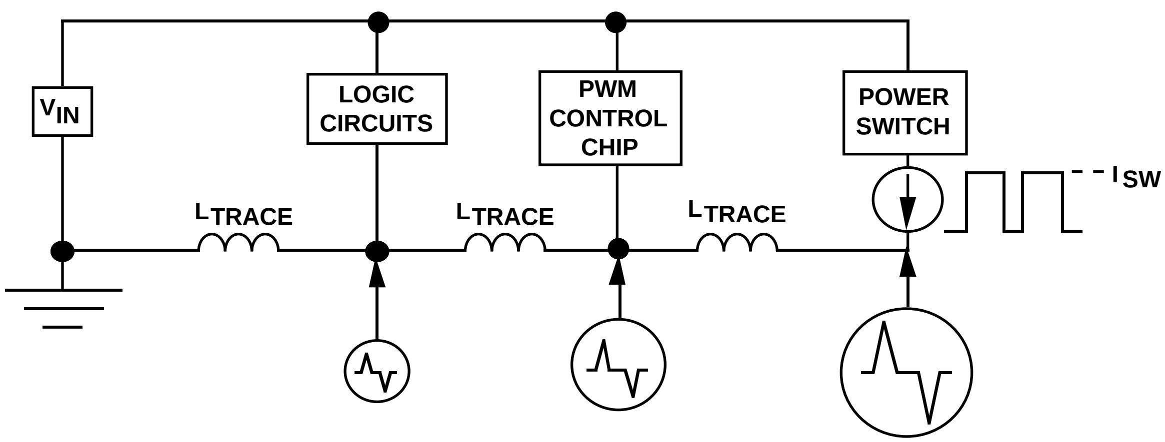


FIGURE 41. EXAMPLE OF POOR GROUNDING

The layout shown has the high-power switch return current passing through a trace that also provides the return for the PWM chip and the logic circuits. The switching current pulses flowing through the trace will cause a voltage spike (positive and negative) to occur as a result of the rising and falling edge of the switch current. This voltage spike follows directly from the **v = L (di/dt)** law of inductance.

It is important to note that the **magnitude of the spike will be different at all points along the trace**, being largest near the power switch. Taking the ground symbol as a point of reference, this shows how all three circuits would be bouncing up and down with respect to ground. More important, **they would also be moving with respect to each other**.

Mis-operation often occurs when sensitive parts of the circuit "rattle" up and down due to ground switching currents. This can induce noise into the reference used to set the output voltage, resulting in excessive output ripple.

Very often, regulators that suffer from ground noise problems appear to be unstable, and break into oscillations as the load current is increased (which increases ground currents).

A much better layout is shown in Figure 42.

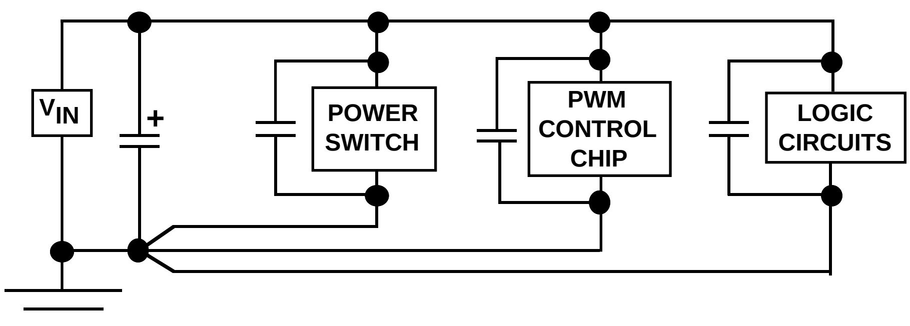


FIGURE 42. EXAMPLE OF GOOD GROUNDING

A big improvement is made by using **single-point grounding**. A good high-frequency electrolytic capacitor (like solid Tantalum) is used near the input voltage source to provide a good ground point.

All of the individual circuit elements are returned to this point **using separate ground traces**. This prevents high current ground pulses from bouncing the logic circuits up and down.

Another important improvement is that **the power switch** (which has the highest ground pin current) **is located as close as possible to the input capacitor**. This minimizes the trace inductance along its ground path.

It should also be pointed out that all of the individual circuit blocks have "local" bypass capacitors tied directly across them. The purpose of this capacitor is RF bypass, so it must be a ceramic or film capacitor (or both).

A good value for bypassing logic devices would be 0.01 µF ceramic capacitor(s), distributed as required.

If the circuit to be bypassed generates large current pulses (like the power switch), **more capacitance is required**. A good choice would be an aluminum electrolytic bypassed with a film and ceramic capacitor. Exact size depends on peak current, but the more capacitance used, the better the result.

Transformer/Inductor Cores and Radiated Noise

The type of core used in an inductor or transformer directly affects its cost, size, and radiated noise characteristics. Electrical noise radiated by a transformer is extremely important, as it may require shielding to prevent erratic operation of sensitive circuits located near the switching regulator.

The most commonly used core types will be presented, listing the advantages and disadvantages of each.

The important consideration in evaluating the electrical noise that an inductor or transformer is likely to generate is the **magnetic flux path**. In Figure 43, the **slug core** and **toroidal core** types are compared.

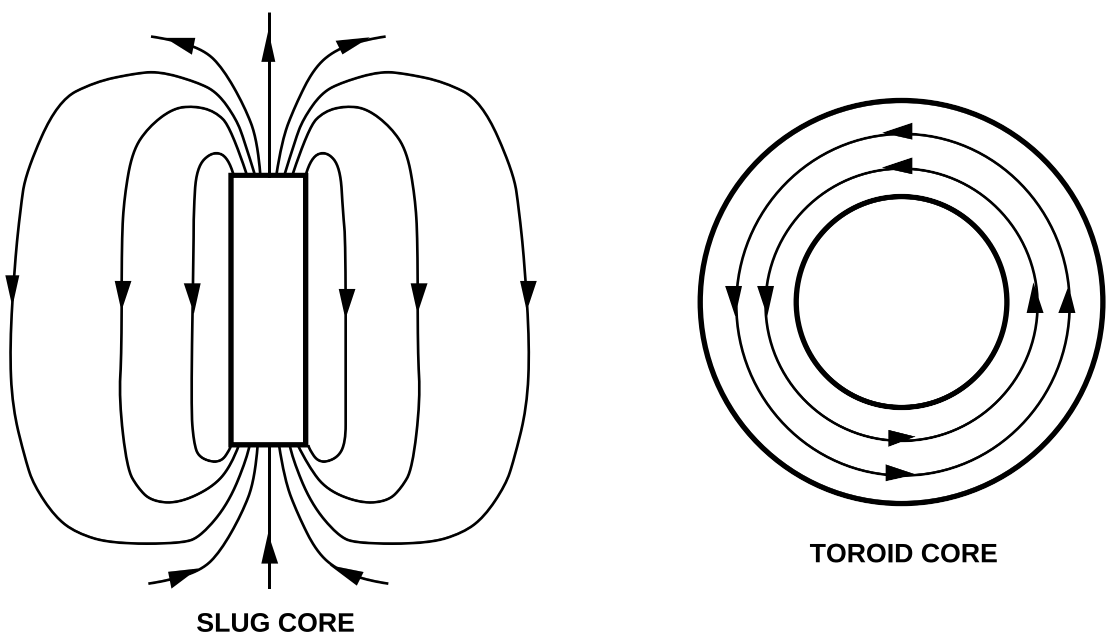


FIGURE 43. FLUX PATHS IN SLUG AND TOROID CORES

The flux in the slug core leaves one end, travels through the air, and returns to the other end. The **slug core is the highest (worst) for radiated flux noise**. In most cases, the slug core device will give the smallest, cheapest component for a given inductor size (it is very cheap to manufacture).

The magnetic flux path in the toroid is **completely contained within the core**. For this reason it has the **lowest (best) radiated flux noise**. Toroid core components are typically larger and more expensive compared to other core types. Winding a toroid is fairly difficult (and requires special equipment), driving up the finished cost of the manufactured transformer.

There is another class of cores commonly used in magnetic design which have radiated flux properties that are much better than the slug core, but not as good as the toroid. These cores are two-piece assemblies, and are assembled by gluing the core pieces together around the bobbin that holds the winding(s).

The cores shown are frequently "gapped" to prevent saturation of the Ferrite core material. The air gap is installed by grinding away a small amount of the core (the gap may be only a few thousandths of an inch).

Figure 44 shows the popular E-I, E-E and Pot cores often used in switching regulator transformers. The cores show the locations where an air gap is placed (if required), but the bobbins/windings are omitted for clarity.

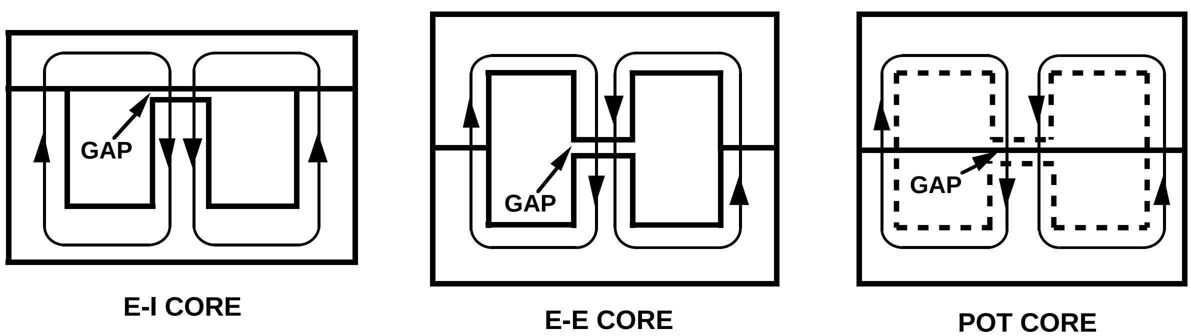


FIGURE 44. FLUX PATHS IN E-I, E-E AND POT CORES

The air gap can emit flux noise because there is a high flux density in the vicinity of the gap, as the flux passing through the core has to jump the air gap to reach the other core piece.

The E-E and E-I cores are fairly cheap and easy to manufacture, and are very common in switching applications up to about 1 kW. There is a wide variety of sizes and shapes available, made from different Ferrite "blends" optimized for excellent switching performance. The radiated flux from this type of core is still reasonably low, and can usually be managed by good board layout.

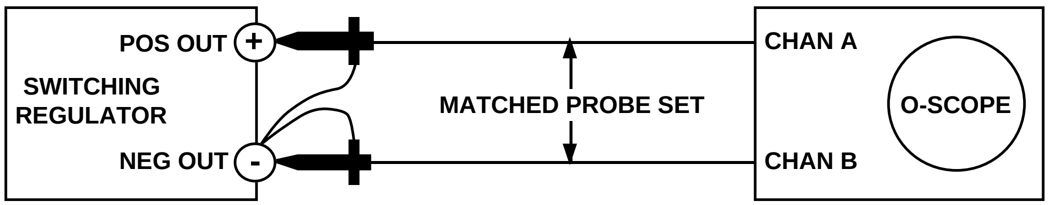
The Pot core (which is difficult to accurately show in a single view drawing), benefits from the shielding effect of the core sides (which are not gapped). This tends to keep the radiated flux contained better than an E-E or E-I core, making the Pot core second best only to the toroid core in minimizing flux noise.

Pot cores are typically more expensive than E-E or E-I cores of comparable power rating, but they have the advantage of being less noisy. Pot core transformers are much easier to manufacture than toroid transformers because the windings are placed on a standard bobbin and then the core is assembled around it.

Measuring Output Ripple Voltage

The ripple appearing on the output of the switching regulator can be important to the circuits under power. Getting an accurate measurement of the output ripple voltage is not always simple.

If the output voltage waveform is measured using an oscilloscope, an accurate result can only be obtained using a **differential** measurement method (see Figure 45).



NOTE: INVERT CHANNEL B AND ADD TO CHANNEL A TO REMOVE COMMON-MODE SIGNAL

FIGURE 45. DIFFERENTIAL OUTPUT RIPPLE MEASUREMENT

The differential measurement shown uses the second channel of the oscilloscope to "cancel out" the signal that is common to both channels (by inverting the B channel signal and adding it to the A channel).

The reason this method must be used is because the fast-switching components in a switching regulator generate voltage spikes that have significant energy at very high frequencies. These signals can be picked up very easily by "antennas" as small as the 3" ground lead on the scope probe.

Assuming the probes are reasonably well matched, the B channel probe will pick up the same radiated signal as the A channel probe, which allows this "common-mode" signal to be eliminated by adding the inverted channel B signal to channel A.

It is often necessary to measure the RMS output ripple voltage, and this is usually done with some type of digital voltmeter. If the reading obtained is to be meaningful, the following must be considered:

1) **The meter must be true-RMS reading**, since the waveforms to be measured are very non-sinusoidal.

2) **The 3dB bandwidth of the meter should be at least 3X the bandwidth of the measured signal** (the output voltage ripple frequency will typically be > 100 kHz).

3) **Subtract the "noise floor" from the measurement**. Connect both meter leads to the negative regulator output and record this value. Move the positive meter lead to positive regulator output and record this value. **The actual RMS ripple voltage is the difference between these two readings**.

Measuring Regulator Efficiency of DC-DC Converters

The **efficiency (η)** of a switching regulator is defined as:

$$\eta = P_{LOAD} / P_{TOTAL}$$

In determining converter efficiency, the first thing that must be measured is the total consumed power (P_{TOTAL}). Assuming a DC input voltage, P_{TOTAL} is defined as the **total power drawn from the source,** which is **equal to:**

$$P_{TOTAL} = V_{IN} \times I_{IN} \text{ (AVE)}$$

It must be noted that the input current value used in the calculation must be the **average value** of the waveform (**the input current will not be DC or sinusoidal**).

Because the total power dissipated must be constant from input to output, P_{TOTAL} is also **equal to the load power plus the internal regulator power losses**:

$$P_{TOTAL} = P_{LOAD} + P_{LOSSES}$$

Measuring (or calculating) the power to the load is very simple, since the output voltage and current are both DC. The load power is found by:

$$P_{LOAD} = V_{OUT} \times I_{LOAD}$$

Measuring the input power drawn from the source is not simple. Although the **input voltage to the regulator is DC, the current drawn at the input of a switching regulator is not**. If a typical "clip-on" current meter is used to measure the input current, the taken data will be essentially meaningless.

The average input current to the regulator can be measured with reasonable accuracy by using a wide-bandwidth current probe connected to an oscilloscope.

The average value of input current can be closely estimated by drawing a horizontal line that divides the waveform in such a way that the **area of the figure above the line will equal the "missing" area below the line** (see Figure 46). In this way, the "average" current shown is equivalent to the value of DC current that would produce the same input power.

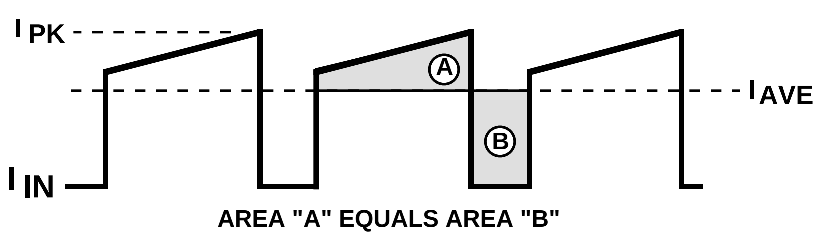


FIGURE 46. AVERAGE VALUE OF TYPICAL INPUT CURRENT WAVEFORM

If more exact measurements are needed, it is possible to **force** the current in the line going to the input of the DC-DC converter to be DC by using an L-C filter between the power source and the input of the converter (see Figure 47).

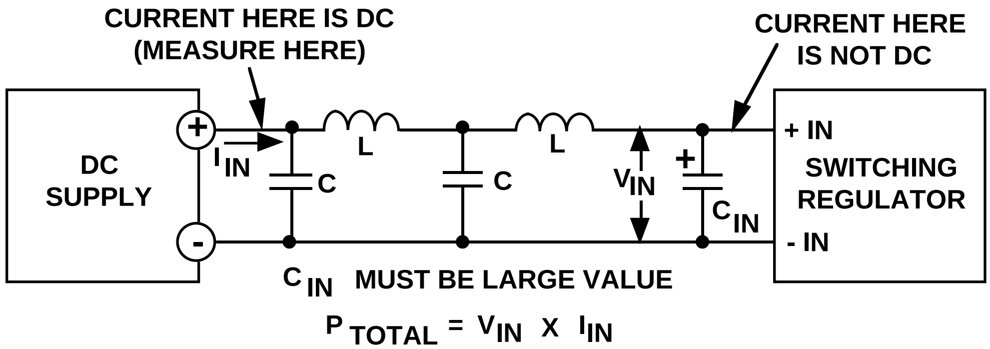


FIGURE 47. L-C FILTER USED IN DC INPUT CURRENT MEASUREMENT

If the L-C filter components are adequate, the current coming from the output of the DC power supply will be DC current (**with no high-frequency switching component**) which means it can be accurately measured with a cheap clip-on ammeter and digital volt meter.

It is essential that a large, low-ESR capacitor be placed at C_{IN} to support the input of the switching converter. The L-C filter that the converter sees looking back into the source presents a high impedance for switching current, which means C_{IN} is necessary to provide the switching current required at the input of the converter.

Measuring Regulator Efficiency of Off-Line Converters

Off-Line converters are powered directly from the AC line, by using a bridge rectifier and capacitive filter to generate an unregulated DC voltage for conversion (see Figures 37 and 38).

Measuring the total power drawn from the AC source is fairly difficult because of the **power factor.** If **both the voltage and current are sinusoidal, power factor is defined as the cosine of the phase angle between the voltage and current waveforms.**

The capacitive-input filter in an off-line converter causes the input current to be **very non-sinusoidal.** The current flows in narrow, high-amplitude pulses (called Haversine pulses) which requires that the power factor be re-defined in such cases.

For capacitive-input filter converters, power factor is defined as:

$$P.F. \quad = \quad P_{REAL} / P_{APPARENT}$$

The **real power drawn from the source** (P_{REAL}) is the power (in **Watts**) which equals the **sum of the load power and regulator internal losses**.

The **apparent power** ($P_{APPARENT}$) is equal to the **RMS input current times the RMS input voltage**. Re-written, the importance of power factor is shown:

$$I_{IN}\,(RMS) = P_{REAL} \,/ \left(V_{IN}\,(RMS) \times PF \right)$$

The RMS input current that the AC line must supply (for a given **real power** in Watts) increases directly as the power factor reduces from unity. Power factor for single-phase AC-powered converters is typically about 0.6. If three-phase power is used, the power factor is about 0.9.

If the efficiency of an off-line converter is to be measured, power analyzers are available which will measure and display input voltage, input power, and power factor. These are fairly expensive, so they may not be available to the designer.

Another method which will give good results is to measure the power **after the rectifier bridge and input capacitor** (where the voltage and current are DC). This method is shown in Figure 48.

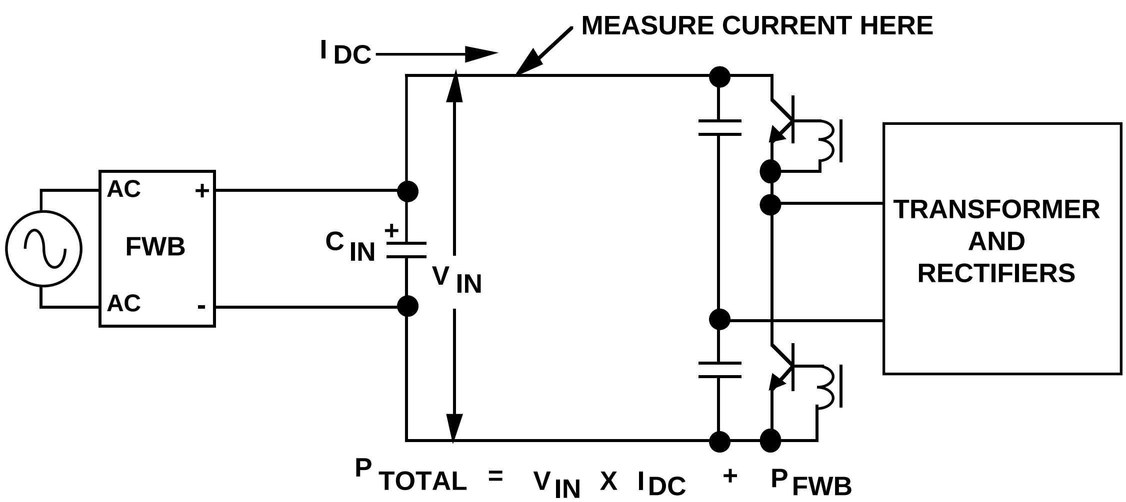


FIGURE 48. MEASURING INPUT POWER IN OFF-LINE CONVERTER

The current flowing from C_{IN} to the converter should be very nearly DC, and the average value can be readily measured or approximated (see previous section).

The total power drawn from the AC source is the sum of the power supplied by C_{IN} (which is V_{IN} X I_{DC}) and the power dissipated in the input bridge rectifier. The power in the bridge rectifier is easily estimated, and is actually negligible in most off-line designs.

Appendix A: Designing the Unregulated Input Supply for a Linear Regulator

Abstract:

The majority of linear regulator applications are powered from unregulated DC sources which derive their power from the AC line.

This paper provides the designer with sufficient information to design a transformer-powered input supply, as well as select the required components (transformer, capacitor, rectifiers).

Transformer theory of operation is provided in sufficient detail to give the reader an understanding of the applicable component specifications.

The circuit topologies covered are the Full-Wave Center-Tapped (FWCT) and the Full-Wave Bridge (FWB) circuits, which account for the vast majority of designs in use.

THE TRANSFORMER POWERED INPUT SUPPLY

In designing an unregulated input supply which is transformer-powered, the objective of the designer is to get the optimum tradeoff for the cost and size of the heatsink, filter capacitor, and transformer. This becomes more critical in higher current supplies, because this can greatly affect size and cost of the overall product.

It is particularly important that the designer know **worst-case values for the minimum and maximum DC input voltage for all operating conditions,** being certain that these voltages are within the allowable limits. If the input voltage falls **too low**, the regulator output will drop out of regulation (and will no longer provide a regulated DC voltage to the load).

If the input voltage goes **too high**, the power dissipation in the regulator can become excessive and cause thermal shutdown. If the input voltage exceeds the absolute maximum rating, the part can be destroyed without drawing **any** load current.

In **all** cases, it is recommended that **any** design be fully tested for high-line and low-line performance and component operating temperatures before it is finalized (transformer turns ratio adjustments are often required to optimize the input voltage range to the regulator).

The Unregulated DC Input Supply

This appendix will address the topologies used in the majority of applications where a transformer is used to generate an unregulated DC voltage (see Figure A-1):

Full Wave Bridge **(FWB)**
Full Wave Center-Tapped **(FWCT)**

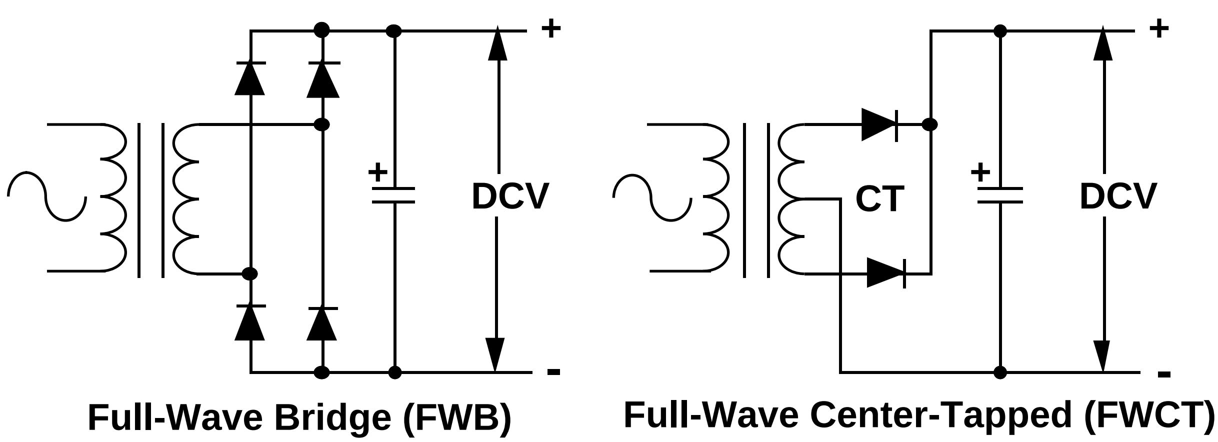


FIGURE A1. TOPOLOGIES FOR DC INPUT SUPPLY

Selecting Full-Wave Bridge vs. Full-Wave Center-Tapped

The best choice for a specific application depends primarily on the load current and output voltage. A very important **difference** between the two circuits is that the **FWB circuit utilizes the full transformer secondary at all times**, while the **FWCT utilizes only one-half of the transformer secondary voltage at any given time** (since only one side of the secondary conducts on each AC voltage half-cycle). This also means a center-tapped transformer secondary connected into a **FWCT** circuit will put out only about **half the DC voltage it would if it was connected to a FWB circuit**.

ADVANTAGES OF FWB CIRCUIT

The FWB configuration is the **best choice for the majority of applications**. The main advantage of this circuit is that the **entire transformer secondary conducts current 100% of the time** (not true of the FWCT). This results in better transformer utilization, and will usually allow a **smaller transformer size to be used in comparison to the transformer which would be required by a FWCT circuit supplying the same load power**.

ADVANTAGES OF FWCT CIRCUIT

The transformer secondaries of the FWCT conduct only 50% of the time. The winding on either side of the center-tap will conduct on alternate half cycles of the AC waveform, which means the secondary winding is not being used as efficiently as in the FWB circuit.

The FWCT may be advantageous in cases where the **output voltage is very low**, or the **load current is very high**. If the regulated output was very low (like 2V or 3V), the added voltage drop across the two additional diodes required in the FWB circuit could make it a less desirable choice (since the additional 1V of transformer secondary voltage required for the extra diode drop could be 30% or 40% of the total secondary voltage).

In cases where the load current is high (like 5A or more), the added power dissipation in the two extra input rectifier diodes can be significant. In some cases it is more cost effective to put a little more money into the transformer to get sufficient additional VA to power the FWCT circuit, thereby avoiding a large heatsink assembly for the input rectifiers.

Understanding Transformer Voltage and Current Ratings

A transformer is usually specified with a primary (input) voltage, a secondary voltage, and a secondary current (load regulation is not typically specified). In order to understand these specifications , a representation of a transformer will be analyzed (see Figure A-2):

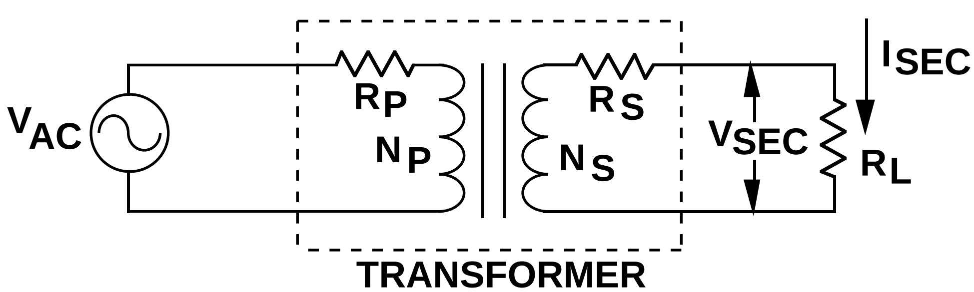


FIGURE A-2. TRANSFORMER MODEL

BASIC TRANSFORMER OPERATION

The transformer **secondary voltage (V_{SEC})** is specified by the manufacturer as an **RMS voltage** which will be measured at the secondary terminals while the **RMS secondary current (I_{SEC})** and **AC input voltage (V_{AC})** are held at their specified values. When tested, the resistive loading (see Figure A-2) means that the **secondary voltage and current are both sinusoidal** (since the input voltage is sinusoidal). When a transformer is connected to a FWB or FWCT circuit with a capacitive filter, **the secondary current is NOT sinusoidal**.

As shown in Figure A-2, the transformer windings have a finite DC resistance (sometimes referred to as **equivalent series resistance**, or simply **ESR**) shown in the figure as R_P and R_S. This resistance is primarily responsible for a characteristic of all transformers called **load regulation**, which is defined as the **percentage change in secondary voltage when the secondary current goes from zero to full rated current**. Load regulation is important in determining the **maximum voltage which will be applied to the regulator input when the load current is light**.

Transformer load regulation is usually worst for small transformers (<20 VA), and better for larger transformers (larger transformers usually use thicker wire for windings which reduces the ESR).

Calculating the Maximum DC Input Voltage

The maximum DC voltage (see Figure A-1) generated by the FWB or FWCT circuit is equal to:

$$V_{DC}\text{ (max)} = V_{SEC}\text{ (RMS)} \times 1.414 \times \frac{VAC_{IN}\text{ (max)}}{VAC_{IN}\text{ (nom)}} \times (1 + \%\text{ Transformer Load Reg.})$$

(Note: for the FWCT circuit, use the voltage of **either separate secondary winding,** which is **half** of the voltage of the total secondary winding).

The maximum secondary voltage of the transformer is NOT the RMS secondary voltage specified on the transformer data sheet. The maximum secondary voltage is the voltage which is measured at the transformer secondary terminals when there is **no load** on the secondary winding (and the **AC input voltage is maximum**). The maximum secondary voltage will be **higher** than the **"loaded" secondary voltage** (the RMS voltage specified by the manufacturer) due to the **transformer load regulation**. The load regulation is usually not specified by the manufacturer, but for small (<20 VA) transformers it can be **30% or more**.

The load regulation effect often causes the absolute maximum ratings for **maximum input voltage** to the regulator IC to be violated. In cases where a 24V output is desired, a regulator like the LM7824 is typically selected. The transformer secondary voltage often selected for a FWB to power the LM7824 is usually about 28V (necessary to provide enough voltage headroom for capacitor ripple, loading effects, and low-line operation).

The **maximum** possible DC voltage applied to the regulator input will be (assuming 30% transformer load regulation and a maximum AC input variation of 10% above nominal):

$$V_{DC}\text{(MAX)} = 28 \times 1.414 \times (1.1) \times (1.3) = \mathbf{57V}$$

It is clear that this design could easily kill a regulator whose maximum allowable input voltage is 40V.

When designing the input DC supply, it is **essential that the maximum input voltage be within the allowable specifications for the linear regulator**. If necessary, a pre-regulator or clamp circuit should be used to protect the linear regulator against damage from input over-voltage conditions (see the **Overvoltage Protection** section later in the paper).

Minimum DC Input Voltage

The minimum DC voltage applied to the input of the linear regulator is critical to assure the regulator does not drop out of regulation. The **lowest** point of the input voltage (**DCV**) **must stay above the regulated output voltage** by an amount that is **equal to or greater than the dropout voltage of the regulator**. Accurately predicting the minimum DC voltage that will result from a given transformer and filter capacitor is not simple, for reasons which will be explained.

A characteristic of capacitive-filtered circuits such as the FWB and FWCT shown in Figure A-1 is that the current flowing from the transformer secondary flows in narrow, high-amplitude pulses which are **not sinusoidal**. The **diode conduction period begins when the rising AC waveform exceeds the voltage level of the capacitor**, and the capacitor begins to charge. The conduction period ends when the falling AC waveform drops below the capacitor voltage, and the capacitor then discharges into the input of the linear regulator until the next rising AC voltage cycle comes along.

An important point to remember is that since the **current flows only in short-duration, high-amplitude pulses, the current flowing through the diodes during the conduction period is much higher than the average current**. This can cause a drop in the peak amplitude of the AC secondary waveform during conduction, as this peak voltage is **reduced** by an amount equal to the **instantaneous current multiplied times the transformer ESR** (it can also cause the top of the AC waveform to be distorted due to loading during the conduction period as illustrated in Figure A-3).

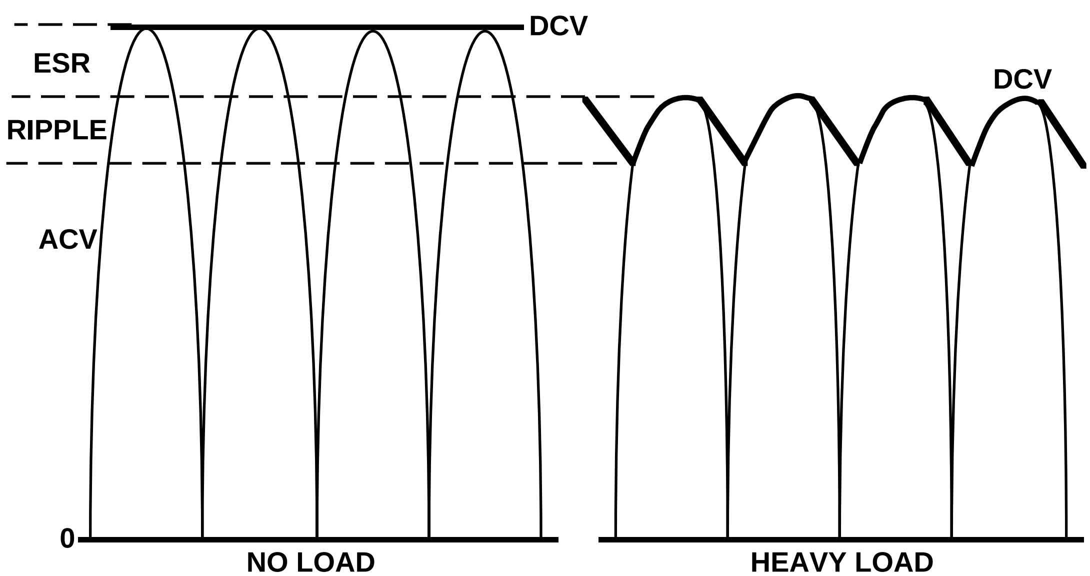


FIGURE A-3: TRANSFORMER LOADING EFFECTS

The second factor which limits the input voltage is the ripple on the filter capacitor. The ripple voltage can be closely approximated by:

$$I = C \times \Delta V / \Delta T$$

Where:

I is the current flowing into the input of the regulator.

C is the value of the capacitor.

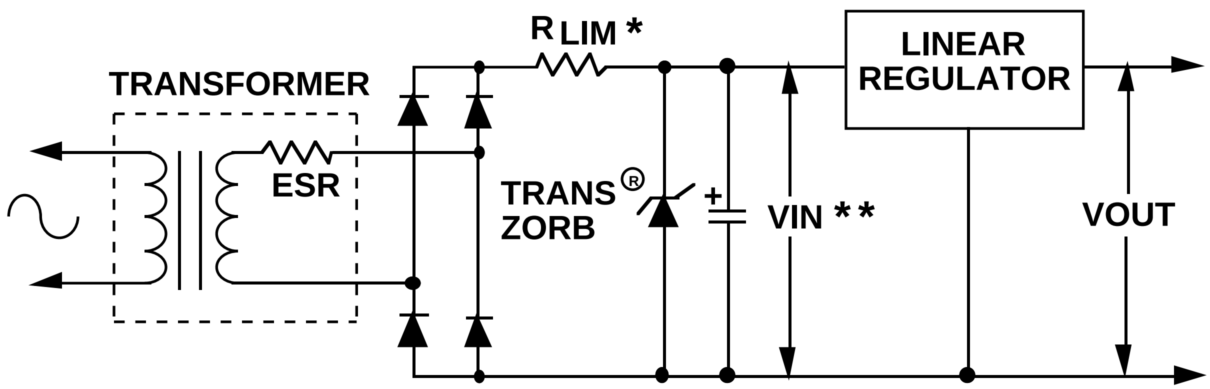
Δ**V** is the peak-to-peak ripple voltage across the capacitor.

Δ**T** is the time period the capacitor discharges (use 6 ms for 60 Hz full-wave rectification)

The maximum ripple that can be allowed in a given application determines the minimum amount of filter capacitance required (as well as capacitor ESR and ripple current rating). Guidelines for component selection are covered later in this paper.

Overvoltage Protection

In cases where the absolute maximum rating for input voltage to a linear regulator may be exceeded under normal operating conditions (worst-case is high-line/no load), some kind of protection must be provided to assure survival of the regulator. A **TransZorb**® (which is equivalent to a power Zener diode) can be used to clamp the maximum input voltage applied to the regulator (see Figure A-4).



**** VIN MUST BE ≤ ABS MAX RATING OF REGULATOR UNDER ALL CONDITIONS**

*** R_{LIM} IS NOT REQUIRED IF ESR IS HIGH ENOUGH TO LIMIT TRANSZORB CURRENT**

FIGURE A-4. INPUT OVERVOLTAGE PROTECTION

The resistor **R_{LIM}** is used to limit the current (and power dissipation) in the clamping element (TransZorb, Zener, etc.) when the input voltage rises high enough to turn it on. It should be noted that **R_{LIM} is effectively in series with the ESR of the transformer**, so the **maximum current through the clamping element is determined by the sum of the two resistances and the peak input voltage**. In some cases (like small, low-current transformers) the ESR may be sufficient to limit the current through the clamping element without additional resistance.

® **TransZorb is a registered trademark of General Semiconductor.**

SELECTING A CLAMPING DIODE

The most important criteria is **clamping voltage**, which is the voltage across the clamp when it is turned on. This information is contained in three specifications, **"Reverse Stand-off Voltage", "Minimum/Maximum Breakdown Voltage", and "Maximum Clamping Voltage"**.

Reverse stand-off voltage is the maximum voltage which the clamping device can tolerate without any significant current flowing into it (the clamp is effectively turned off). This tells where the clamp is **off,** but the most important value is where it turns **on**.

Minimum/maximum breakdown voltage is the voltage range where the clamping device begins to turn on (and clamp). This gives information on where clamping action **begins**, but does not tell how high the clamped voltage can rise.

Maximum clamping voltage gives the **highest** voltage that can be seen across the clamping element **at a specified value of current**. This is an important specification, since it defines the maximum voltage which can appear at the regulator input.

Selecting Components for the Unregulated DC Input Supply

This section gives information which enables the user to select the appropriate transformer, diodes, and capacitors to use in the unregulated DC supply powered from a transformer in either the FWB or FWCT configurations (see Figure A-1).

SELECTING THE TRANSFORMER

In selecting the appropriate transformer to use in the unregulated input supply, it will be assumed the following specifications are known:

AC Input Line Voltage and Frequency
Regulated DC Output Voltage
DC Output Current Maximum
Topology (FWB or FWCT)

The transformer specifications which must be determined are:

Secondary RMS Voltage Rating
Secondary RMS Current Rating

<u>Secondary Current Rating</u>

The RMS secondary **current** rating required by a transformer powering a FWB and FWCT circuit can be closely approximated using:

For FWB: I_{SEC} = 1.8 X Max. DC Output Current
For FWCT: I_{SEC} = 1.2 X Max. DC Output Current

It might appear at first inspection that the FWCT circuit is better, because it requires a secondary winding with a **lower current rating** (which might mean a smaller overall transformer size). However, it must be remembered that the voltage rating of the secondary winding of a transformer used with a FWCT must be approximately **twice** the voltage rating of a transformer used with a FWB to get the same unregulated DC voltage. This means the **Volt-Amp** (VA) rating of the transformer (which is the product of the secondary voltage and current ratings), must be about **33% larger** if used to power the FWCT circuit (and it is the VA rating that most accurately reflects transformer size and weight). This shows the main advantage of using the FWB circuit, as it **minimizes** the size of the transformer required for a given DC output voltage and current.

Secondary Voltage Rating

The RMS secondary **voltage** rating required for a transformer can be approximated by using:

$$V_{AC}\text{ (RMS SEC)} = \frac{V_{OUT} + V_{REG} + V_{RECT} + V_{RIPPLE}\text{ (pk)}}{0.9} \times \frac{V_{NOM}}{V_{LOWLINE}} \times 0.707$$

Where the terms in the equation are defined as:

$\mathbf{V_{OUT}}$ is the DC output voltage of the regulated output.

$\mathbf{V_{REG}}$ is the minimum voltage drop required across the linear regulator to assure regulation (this can be assumed to be about **3V** for standard regulators, and about **1V** for most LDO type regulators).

$\mathbf{V_{RECT}}$ is the total voltage drop across the rectifier diodes. For silicon rectifiers, assume a diode drop of about 1V (in the FWCT only **one** diode conducts at a time, but in the FWB there are always **two** diodes conducting).

FWCT: $V_{RECT} = 1V$ **FWB**: $V_{RECT} = 2V$

Note: If **Schottky** rectifiers are used, V_{RECT} values of **0.5V** and **1V** can be used for FWCT and FWB respectively.

$\mathbf{V_{RIPPLE}}$ is the ripple voltage across the input filter capacitor. The value entered into the equation should be in **Volts peak**, which is **one-half of the peak-to-peak value** which can be calculated from the approximation:

$$V_{RIPPLE}\text{ (pk-pk)} = \frac{I_{LOAD}}{\text{Capacitance}} \times (.006)$$

This highlights an important trade-off: Reducing the ripple voltage allows a lower transformer secondary voltage to be used (which means a smaller transformer), but the cost is increased capacitor size. As a rule of thumb, a capacitor should be selected which allows a **maximum peak ripple voltage which is about 10% of the DC output voltage** (which means the **peak-to-peak ripple would be 20% of the DC output voltage**).

V_{NOM} is the nominal AC input line voltage applied to the transformer primary. This is typically 115V, 230V, or 208V for most of the power used in the USA.

$V_{LOWLINE}$ is the minimum AC input voltage which will be applied to the transformer primary. In most commercial applications, AC line droop of 10% to 15% must be tolerated by the power supply without loss of performance.

FWB vs. FWCT

IMPORTANT: When calculating the secondary voltage for the transformer, remember to differentiate between the two topologies:

FWCT: In this case only one diode drop is used in the calculation of V_{RECT}. When a value for the secondary voltage is obtained, remember the **FWCT requires a center-tapped secondary whose voltage rating is twice the calculated value**.

FWB: The FWB uses two diode drops in the calculation of V_{RECT}. The value calculated for secondary voltage is equal to the transformer secondary voltage rating, and a center-tap is not used.

SELECTING THE INPUT FILTER CAPACITOR

When selecting the input filter capacitor to be used with the unregulated DC supply, the most important specifications which must be defined are:

- Voltage Rating
- Amount of Capacitance Required
- Ripple Current Rating

Input Capacitor Voltage Rating

The voltage rating on the input filter capacitor must be **greater than the maximum peak voltage that will be applied to it under any operating condition**. The maximum value will occur at high AC line, with minimum load current. The peak DC voltage can be calculated using the following equation:

$$V_{DC}\text{ (max)} = V_{SEC}\text{ (RMS)} \times 1.414 \times \frac{VAC_{IN}\text{ (max)}}{VAC_{IN}\text{ (nom)}} \times (1 + \%\text{ Transformer Load Reg.})$$

The terms in the equation are defined as:

V_{SEC} (RMS) is the specified transformer secondary voltage at full load current (this is the specification provided on the transformer data sheet).

VAC_{IN} (max) is the maximum AC input voltage that will be applied to the transformer primary.

VAC $_{IN}$ (nom) is the nominal value for the AC input voltage used to power the transformer primary.

% Transformer Load Regulation is defined as the percentage change in AC secondary voltage that is measured when the secondary current goes from full rated RMS current to no current (this topic was discussed previously in this paper under **Basic Transformer Operation**). This specification is not typically provided on the data sheet for a transformer, so it must be measured by actual testing or estimated. For transformers rated **10 VA or less, assume about 30%.** Transformers with a higher VA rating will have a better (lower) load regulation figure, but the best information will probably be obtained from bench testing (or from the manufacturer).

Amount of Capacitance Required

The amount of capacitance required is determined by the **load current and the maximum allowable ripple voltage**. The capacitance can be calculated using the formula:

$$\text{Capacitance} = \frac{I_{LOAD}}{V_{RIPPLE\ (pk\text{-}pk)}} \times (.006)$$

As a good rule of thumb, the capacitor should be selected so that the **peak-to-peak ripple voltage is about 20% of the regulated DC output at maximum load current** (which is the maximum DC current delivered by the regulated output).

Ripple Current Requirement

The RMS ripple **current** that flows into the filter capacitor is **NOT** equal to the DC load current (it is actually about 2 or 3 times this value). This ripple current flows through the internal resistance of the capacitor and generates internal heat. In lower current DC supplies (like less than 0.5A), ripple current is not usually a problem. In higher current supplies, ripple current sometimes causes input capacitor failure due to overheating. The designer must make certain that the filter capacitor selected can support the ripple current without becoming hot enough to reduce reliability, which sometimes necessitates using a **larger** capacitor than would be required by ripple **voltage** calculations alone.

SELECTING THE RECTIFIER DIODES

The specifications which must be defined for the rectifier diodes are:

Blocking Voltage
Average Current
Surge Current

Blocking Voltage

The blocking voltage rating of a diode tells how much reverse voltage the diode can safely tolerate when it is not conducting. The voltage rating required for a given application is dependent on the transformer secondary voltage (and load regulation), as well as the type of circuit (FWB or FWCT).

FWB: First calculate the maximum DC voltage which will appear across the filter capacitor (see Input Capacitor Voltage Rating) for high-line/no-load conditions. The **diode blocking voltage for FWB circuits must be greater than or equal to the maximum DC voltage calculated**.

FWCT: Calculate the maximum DC voltage which will appear across the filter capacitor (see Input Capacitor Voltage Rating) for high-line/no-load conditions. The **diode blocking voltage for FWB circuits must be greater than or equal to twice the maximum DC voltage calculated**.

Average Current

The average current specification tells how much current the diode can safely conduct continuously. In either the FWB or FWCT circuits, the diodes conduct only half the time (on each alternate AC half-cycle). Remember also, **the average current through the secondary winding (and the diodes) is NOT equal to the load current** (as a rule of thumb for FWCT, $I_{AVE} = 1.2\ I_{LOAD}$ and for FWB, $I_{AVE} = 1.8\ I_{LOAD}$). Remembering the 50% duty cycle of the diodes, the **minimum average current rating** necessary for the rectifier diodes can be calculated from:

$$\text{FWCT:}\quad I_{AVE} = 0.5 \times (1.2\ I_{LOAD})$$

$$\text{FWB:}\quad I_{AVE} = 0.5 \times (1.8\ I_{LOAD})$$

Surge Current

When power is first applied to a FWB or FWCT circuit (and the voltage on the input capacitor is zero), a large surge current flows from the secondary winding, through the rectifier diodes, and into the capacitor. A good approximation for the maximum (worst-case) value of this surge current can be found by:

$$I_{SURGE}\,(\text{max}) = \frac{V(\text{RMS})_{SEC} \times 1.414}{\text{Impedance}}$$

Where the term Impedance equals the **sum of the resistance of the secondary winding** (which can be measured with an Ohm meter) **added to the ESR of the input filter capacitor** (which is specified on the capacitor's data sheet).

Surge current ratings for diodes are typically many times higher than average current ratings (for 1N4001, the average current rating is 1A, while the surge current rating is 30A). For most low-current (< 500 mA) transformers, the DC resistance of the secondary winding is high enough so that surge current will not damage the rectifiers typically used.

High-current transformers (which have thick wire and low winding ESR) are often used with very large input filter capacitors (which also have very low ESR). Surge current **must be considered** when selecting the diodes used in such an application.

Appendix B: Selecting the Correct Heatsink to use with an IC Regulator

Abstract:

There are many applications of IC regulators where the part must dissipate significant amounts of power, which requires a heatsink to keep the operating temperature of the part within safe limits.

This paper presents methods to calculate the dissipated power in a specific application, and select the appropriate heatsink for a reliable design.

The focus is on the most commonly used power packages, the **T0-220** and **T0-3**.

Theory on thermal resistance and heat transfer is presented in sufficient detail to give the designer an understanding of heatsink specifications.

Heatsink Specification and Selection

POWER DISSIPATION IN A LINEAR REGULATOR

All semiconductors dissipate power when they operate, and this power dissipation results in an increase in the junction temperature of the part.

The critical design consideration is to make certain that the heat resulting from the regulator's internal power dissipation is conducted away from the die quickly enough to prevent the die temperature from exceeding safe limits (in most cases, 125 °C is an appropriate maximum value for reliable designs).

Heat is conducted away from the regulator using a **heatsink,** which must be selected based on the power dissipation in a specific application.

CALCULATING POWER DISSIPATION

Power dissipation in any device is given by the relation:

$$P = \Delta V \times I$$

Where:

P is the power dissipated in Watts.

ΔV is the voltage potential difference in Volts.

I is the current in Amps.

Figure B-1 shows the equivalent circuit for a linear regulator which will be used to explain how to calculate power dissipation:

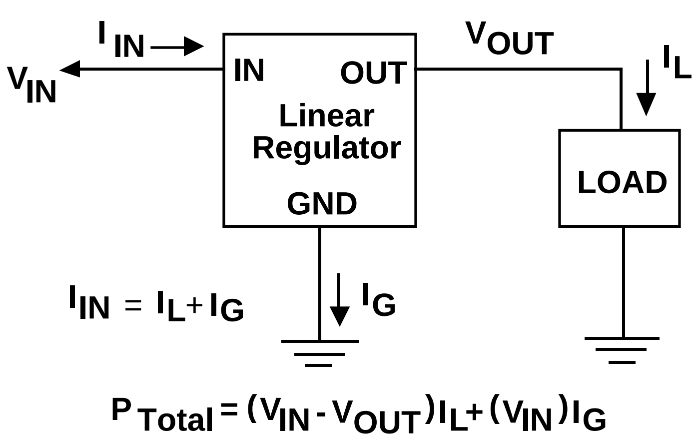


FIGURE B-1: POWER DISSIPATED IN A LINEAR REGULATOR

The equation shown in Figure B-1 which is used to calculate the total power dissipation contains two terms (note: the **total** power dissipation in **any** linear regulator application is found by **adding these two terms together)**:

LOAD CURRENT TERM: The current which flows into the regulator input and then out the output pin (and into the load) generates power dissipation as given by:

$$P_L = I_L \times (V_{IN} - V_{OUT})$$

GROUND PIN CURRENT TERM: The current which flows into the regulator input and then out the ground pin back to the source generates power dissipation as given by:

$$P_G = I_G \times V_{IN}$$

The ground pin current in most **standard (NPN Darlington)** type regulators is low enough that the power dissipation resulting from it is negligible compared to the load current term. In **low-dropout (LDO)** regulators, the ground pin current term may be significant in contributing to the total power dissipation.

DETERMINING VALUES FOR POWER DISSIPATION CALCULATIONS

When using the equations shown to calculate power dissipation, care must be taken to be sure that worst-case values are used (the values which will generate the **maximum** power dissipation under any anticipated operating conditions). Using worst-case values guarantees the heatsink selected will be adequate for the application.

LOAD CURRENT: The load current value used in calculations must be the **maximum** expected load current (not the **nominal** value) which can occur under normal operation.

INPUT VOLTAGE: Systems which operate from an AC line often must be specified to work with a **line tolerance of +/-10% or even +/-20%.** The **AC high-line value must be used to calculate or measure the input voltage** (since this will give the maximum value for input voltage and maximum power dissipation).

Note: The input voltage derived from a rectified AC source will have an AC ripple component as shown in Figure B-2. The value used for V_{IN} in the power dissipation calculations should be the **average** value.

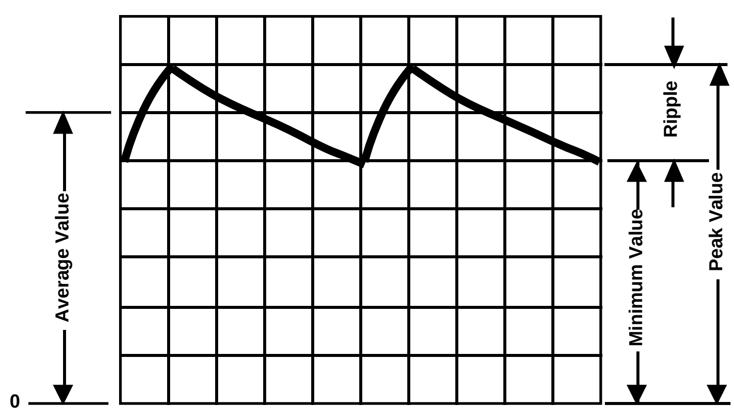


FIGURE B-2. TYPICAL INPUT VOLTAGE WITH RIPPLE COMPONENT

HEAT TRANSFER AND HEATSINKING

The power dissipated within a linear regulator causes the die temperature to rise, which means this heat must be conducted away from the die to prevent the part from overheating. Fortunately, one of the laws of thermodynamics says that heat will always flow from the hotter body (the regulator dissipating the power) to a cooler body (usually the surrounding air) if you give the heat some path of conduction.

The designer must be certain that the **thermal resistance** (the resistance to heat flow) of the heat's conduction path is low enough that the heat will flow out of the part fast enough to keep it from overheating. This is fairly simple to calculate once the power dissipation and ambient temperature are known.

THERMAL RESISTANCE

Thermal resistance, symbolized by θ, can be thought of as a measure of how well a material prevents heat from flowing through it. **An ideal conductor of heat would be one whose resistance is zero** (which means better heatsinking occurs with **lower** values of thermal resistance).

A simple circuit will be used to explain heat flow (see Figure B-3).

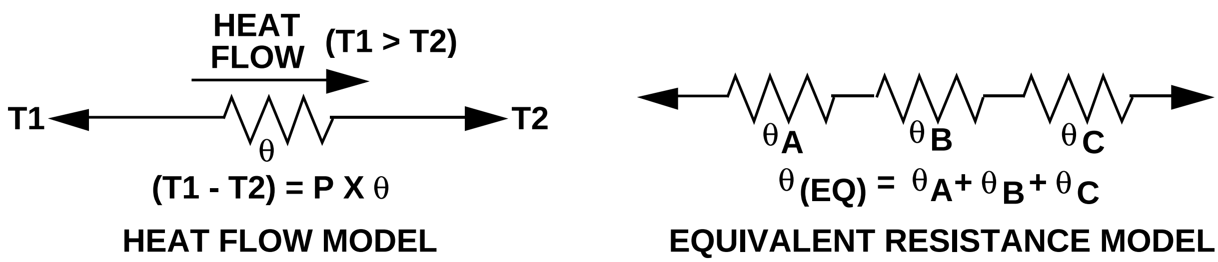


FIGURE B-3. THERMAL CIRCUIT MODELS

Figure B-3 shows a model of a thermal circuit showing heat flow, thermal resistance, and temperature differential. In the thermal circuit, it can be seen that heat will always try to flow from the hotter point to the cooler point through the path which is represented by the thermal resistance (and the lower the resistance, the easier the heat can flow).

It is important to note that the equivalent of a **series** of resistances is found by simply **adding together each of the individual resistances**.

The unit of measure for thermal resistance is °**C/W**. The term °C/W is convenient for specifying and selecting a heatsink, since the it allows the designer to specify (select) a heatsink once the power dissipation and ambient temperature are known (this will be covered in more detail later).

Obtaining an accurate number for thermal resistance requires evaluating all of the thermal interfaces that the heat must cross in going from the die to the ambient (this is covered in the next section).

HOW TO CALCULATE THERMAL RESISTANCE

The total thermal resistance in any application can only be found by adding together all of the individual thermal resistances through which the heat must flow. This is illustrated by a diagram showing the top view of a typical TO-220 linear regulator attached to a heatsink (see Figure B-4):

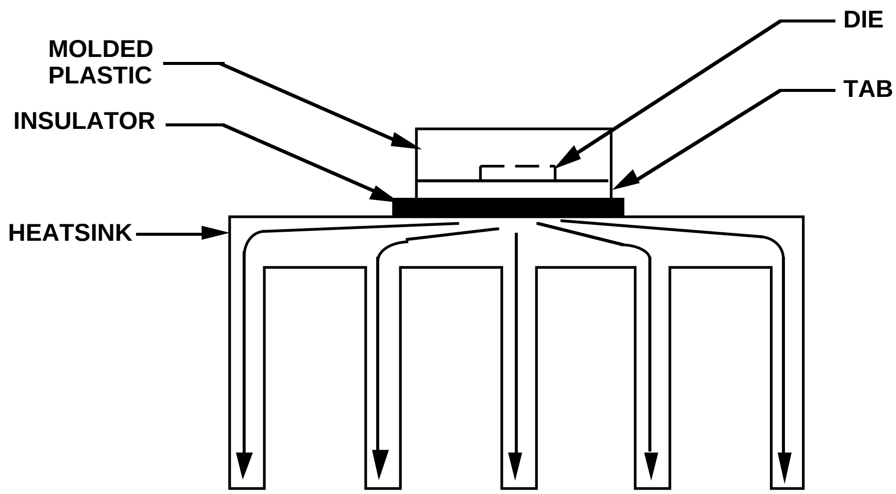


FIGURE B-4. THERMAL RESISTANCE MODEL

In looking at the diagram, it can be seen that the heat starts at the point of power dissipation (the die) and flows through the metal tab, the insulator, and into the heat sink. The heat is then transferred from the heatsink to the surrounding air as air passes through the fins.

This shows that for this example, the heat passes through **three** distinct thermal resistances in reaching the ambient air:

Junction-to-case: The thermal resistance between the die and the surface of the tab (or case).

Case-to-heatsink: The thermal resistance between the tab (case) and the heatsink surface.

Heatsink-to-ambient: The thermal resistance between the heatsink and the ambient air.

<u>Junction-to-case Thermal Resistance</u>

The thermal resistance between the junction (die) and the case (called $\theta_{J\text{-}C}$) is specified on the data sheet for most power products such as linear voltage regulators. It is dependent on die size, package style (TO-3, TO-220, etc.), and die-attach method used. Typical values for parts in a **TO-3 package is about 2 to 3 °C/W, and a TO-220 package would be about 3 to 5 °C/W** (exact values will be listed on the product data sheet)**.**

NOTE: Parts in small packages (like TO-92, DIP, and surface mount) often do not have values for junction-to-case thermal resistance. These parts can not handle much power dissipation (typically about one or two Watts maximum) and can not be easily mounted to a conventional heatsink. The "heatsinking" for these parts is usually made using the copper on the PC board, so they are supplied with data on the **total junction-to-ambient thermal resistance** for typical PC board mounting methods.

Case-to-heatsink Thermal Resistance

The thermal resistance between the tab (case) of the regulator and the heatsink it is attached to (called $\theta_{\mathbf{C\text{-}S}}$) is primarily dependent upon the package type and the insulating material (if used) between the case and the heatsink. Insulators are required in any application where the case must be electrically isolated from the heatsink. There are many materials to choose from such as mica, Kapton, silicon-filled rubber and beryllium oxide.

For example, a TO-3 type case attached to an aluminum heatsink using a mica insulating pad and thermal grease can have a case-to-heatsink thermal resistance as low as **0.3 or 0.4 °C/W**. A similarly mounted TO-220 device would have a case-to-heatsink thermal resistance of about **1.5 °C/W**. This difference gives the TO-3 a big advantage in applications of high power dissipation .

Heatsink-to-ambient Thermal Resistance

$\theta_{\mathbf{S\text{-}A}}$ (defined as the thermal resistance between the point on the heatsink where the heat source is mounted and the surrounding air) is the measurement of how good the heatsink is. This maximum allowable value of $\theta_{\mathbf{S\text{-}A}}$ for a given application (which must be calculated by the designer based upon power dissipation and ambient temperatures) is used to select the heatsink.

Lower values of thermal resistance (which mean better heatsinking) require a larger, heavier heatsink (since it must have more surface area where heat exchange can take place between the heatsink and the surrounding air). For approximate numerical values of $\theta_{\mathbf{S\text{-}A}}$ **(assuming still air)**:

Thermal resistances between **20 and 30 °C/W** are easy to obtain with very small heatsinks (a TO-3 device with **no heatsink at all** will give a **junction-to-ambient** thermal resistance of about 35 °C/W).

Thermal resistances between **10 and 20 °C/W** are still fairly small (requiring about 2 in.^{3} of volume).

Thermal resistances **below 10 °C/W** require fairly large heatsinks. Values as low as **1 °C/W** are available, but would weigh several pounds (or more) and occupy at least 30 or 40 in.^{3} of volume.

Note: increased airflow can also reduce the effective value of $\theta_{\mathbf{S\text{-}A}}$, but requires the addition of a fan.

TOTAL THERMAL RESISTANCE

It must be stressed that calculating the **total thermal resistance from junction-to-ambient requires adding together all of the individual thermal resistances that are present in the specific application (junction-to-case, case-to-heatsink, heatsink-to-ambient)**.

Since all of the resistances add in series, it means that all have equal impact on the total thermal resistance. For example, improving (reducing) the thermal resistance of a heatsink from 5 °C/W to 3 °C/W (a savings of 2 °C/W) will greatly increase the size and cost of the heatsink. However, a similar savings may be obtained by selecting a part in a different package:

The LM350 linear regulator is offered in both the TO-220 and the TO-3 package, with the TO-220 being less expensive. However, in comparing the thermal resistances (junction-to-case)of the two packages, the TO-3 is about 2 °C/W better. If mica insulators are used to mount the parts, an additional 1 °C/W (**for a total of 3°C/W**) improvement is obtained by using the TO-3 instead of the TO-220 package.

DEFINING THE HEATSINK

The heatsink is defined once the following information is known:

T_A (max): the maximum ambient temperature

T_J (max): the maximum junction temperature

P_T: the total maximum power dissipation

$\theta_{J\text{-}C}$: the thermal resistance from junction to case (listed on the data sheet)

$\theta_{C\text{-}S}$: the thermal resistance from case to heatsink (depends on type of insulator used)

The only parameter (which must be calculated) which defines the heatsink is the **thermal resistance from heatsink to ambient**, usually referred to as $\theta_{S\text{-}A}$.

The first step in finding $\theta_{S\text{-}A}$ is to calculate the thermal resistance from junction to ambient $\theta_{J\text{-}A}$:

$$\theta_{J\text{-}A} = \left(T_J\,(\text{max}) - T_A\,(\text{max})\right) / P_T$$

Remembering that the total (junction to ambient) thermal resistance is equal to the sum of the individual resistances:

$$\theta_{J\text{-}A} = \theta_{J\text{-}C} + \theta_{C\text{-}S} + \theta_{S\text{-}A}$$

Rearranging the equation, the thermal resistance from heatsink to ambient $\theta_{S\text{-}A}$ can be isolated:

$$\theta_{S\text{-}A} = \theta_{J\text{-}A} - \theta_{J\text{-}C} - \theta_{C\text{-}S}$$

We can now solve for $\theta_{S\text{-}A}$, since all of the terms on the right-hand side of the equation are known.

SELECTING THE HEATSINK

It is the value of heatsink to ambient thermal resistance $\theta_{S\text{-}A}$ that defines the heatsink required for a given application. This value is sometimes specified directly by the manufacturer on the heat sink data sheet, but often it must be calculated from a graph which shows **temperature rise** plotted versus **power dissipation**. If such a graph is provided, the thermal resistance of the heatsink is found by dividing the temperature rise by the power dissipation at the appropriate point on the curve.

It should be noted that once the value for $\theta_{S\text{-}A}$ is calculated for an application, any heatsink can be used which has a value of $\theta_{S\text{-}A}$ which is **less than or equal to this number**.