AND8093/D

Current Sensing Power MOSFETs

SENSEFET™ PRODUCT

Current sensing power MOSFETs provide a highly effective way of measuring load current in power conditioning circuits. Conceptually simple in nature, these devices split load current into power and sense components, and thereby allow signal level resistors to be used for sampling. Since this technique results in higher efficiency and lower costs than competing alternatives, understanding how to use SENSEFET product is an important design issue.

Getting accustomed to these devices is relatively, but not completely, straightforward. They are conceptually simple, but have their own unique set of characteristics and subtle properties. The following discussion examines both, and starts with a description of how SENSEFET devices work.

Principle of Operation

Their operation is based on the matched devices principle that is so commonly used in integrated circuits. Like integrated circuit transistors, the on–resistance of individual source cells in a power MOSFET tends to be well matched. Therefore, if several out of several thousand cells are connected to a separate sense pin, a ratio between sense section on–resistance and power section on–resistance is developed. Then, when the SENSEFET device is turned on, current flow splits inversely with respect to the two resistances, and a ratio between sense current and source current is established.

The separate source connection is called a mirror. Typically SENSEFET product is designed such that the ratio between mirror cells and source cells is on the order of 1:1000 Schematically, this looks like two parallel FETs with common gate and drain connections, but separate source leads. An illustration of this configuration appears in Figure 1. The relative size of the two devices determines how current is split between source and mirror terminals. The ratio of source current to mirror current is specified by n, the "Current Mirror Ratio." This ratio is defined for conditions where both source and mirror terminals are held

at the same potential. Since n is on the order of 1000:1, load current is approximately equal to source current, and the current mirror ratio also describes the ratio of load current to sense current.
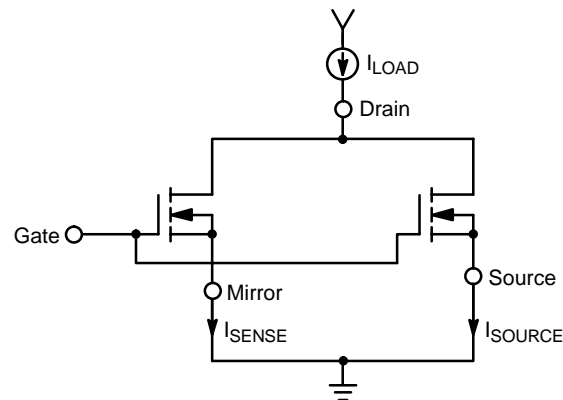

Figure 1. SENSEFET Equivalent Circuit

When a signal level resistor is connected between mirror and source terminals, a known fraction of load current is sampled without the insertion loss that is associated with power sense resistors. For this reason, the technique of measuring load current with SENSEFET devices is called "lossless current sensing." As long as the sense resistor is less than 10% of the mirror section's on–resistance $R_{DM(on)}$, the current that is sampled is approximately load current divided by the current mirror ratio or I_{LOAD}/n. In practice, the amount of sense voltage that is developed with such low values of sense resistance is usually not sufficient to drive current limiting circuits. Therefore, larger values of R_{SENSE} are normally used. These larger values appreciably affect the total resistance in the mirror leg, and therefore, alter the current mirror ratio. How to model this behavior and calculate sensing parameters is discussed as follows.

Calculating Sense Resistance

With the aid of the model that is shown in Figure 2, calculating sense voltage and sense resistance is very straightforward. In this model, $R_{DS(on)}$ is separated into bulk and active components. Bulk drain resistance is common to the entire device, and is represented by R_b. Active components of $R_{DS(on)}$ are modeled by $R_{a(on)}$ for the

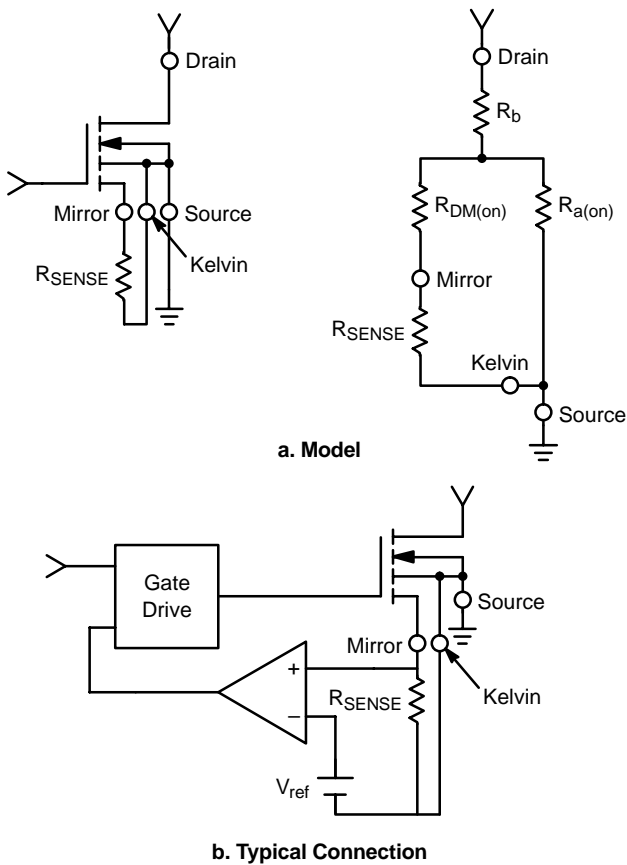**a. Model**

b. Typical Connection

Figure 2. Model and Typical Connection

power section, and $R_{DM(on)}$ for the mirror. R_{SENSE} is the external sense resistor.

If R_{SENSE} is an open circuit, the maximum voltage that can appear at the mirror terminal is $V_{DS(on)} \times R_{a(on)}/(R_{a(on)} + R_b)$. In other words, the mirror terminal does not sample the full drain–source on voltage, but sees only the fraction of drain–source voltage that is represented by $R_{a(on)}/(R_{a(on)} + R_b)$. This ratio is called the mirror compliance ratio, K_{MC}. Values for $R_{a(on)}$ and R_b are determined by measuring the mirror compliance ratio, and multiplying $R_{DS(on)}$ by this ratio to get $R_{a(on)}$. Bulk resistance, R_b, is then determined by subtracting $R_{a(on)}$ from $R_{DS(on)}$. Given these values, $R_{DM(on)}$ is determined by multiplying $R_{a(on)}$ times the current mirror ratio, n.

Given values for the model's internal resistors, sense voltage, sense resistance, and drain current can be calculated from simple resistive divider equations. These equations are summarized as follows:

Sensing Equations

$$V_{SENSE} = I_D \cdot R_{a(on)} \cdot \frac{R_{SENSE}}{(R_{SENSE} + R_{DM(on)})} \quad (1)$$

$$R_{SENSE} = V_{SENSE} \cdot \frac{R_{DM(on)}}{[(I_D \cdot R_{a(on)}) - V_{SENSE}]} \quad (2)$$

$$I_D = V_{SENSE} \cdot \frac{(R_{SENSE} + R_{DM(on)})}{(R_{a(on)} \cdot R_{SENSE})} \quad (3)$$

The results obtained from using these equations agree very well with measured values. Using the MTP10N10M as an example, calculated and measured values are compared in Table 1. They are based upon 5 amps of drain current, $R_{a(on)}$ = 116 milliohms, R_b = 44 milliohms, and $R_{DM(on)}$ = 209 ohms.

Table 1. Calculated versus Measured Sense Voltage

R_{SENSE} Ω	Calculated V_{SENSE} mV	Measured V_{SENSE} mV	Δ %
20	51	50	2
47	106	105	1
100	179	185	−3
200	284	290	−2
1.0 k	480	480	0

Since all of the actual values were measured on an oscilloscope, the discrepancies which are shown here are all within measurement accuracy. Given static conditions, the model in Figure 2 does a good job.

In a typical application such as the one shown in Figure 2b, a current trip is produced when V_{SENSE} is equal to the comparator's reference voltage, V_{ref}. Therefore, substituting V_{ref} for V_{SENSE} in these equations yields combinations of I_D and R_{SENSE} for which a current limit signal is produced.

For reasons which will soon be discussed, it is generally advisable to choose a value of R_{SENSE} that does not exceed $R_{DM(on)}$. As the values in Table 1 indicate, this constraint produces sense voltages on the order of 250 mV at the MTP10N10M's normal operating current. Although this is sufficient for most applications, lower operating currents and device types with lower mirror compliance ratios can lead to problems with generating usable values of sense voltage. Where higher values of sense voltage are required, the technique shown in Figure 3 can be used. In this circuit,the SENSEFET mirror is held at the same potential as its source, and op amp OA1 generates a negative output voltage that equals sense current times the feedback resistor R_f. Sensing equations for this type of virtual ground circuit are listed as follows:

Virtual Ground Sensing Equations

$$V_{SENSE} = -I_D \cdot n \cdot R_f \quad (4)$$

$$R_f = V_{SENSE}/I_D \cdot n \quad (5)$$

$$I_D = -(V_{SENSE}/R_f)n \quad (6)$$

These equations assume that the op amp's input bias current and input offset voltage are both zero. Using some of today's newer op amps, this assumption is a good one. With an MC34081 for example, room temperature values for input bias current and input offset voltage are less than one nanoamp and less than one millivolt, respectively.
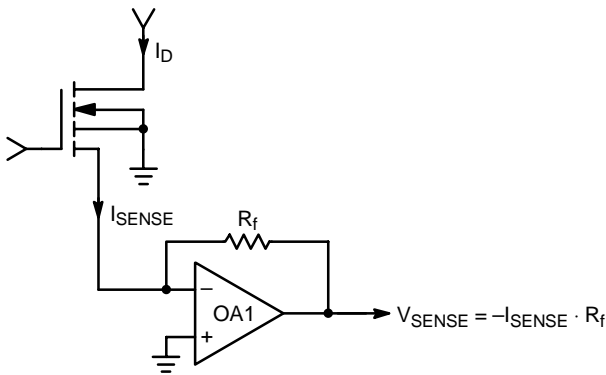
Figure 3. Virtual Ground Sensing

Although virtual ground sensing does a nice job of boosting signal level, it requires a negative power supply and produces a negative output signal, both of which can be undesirable. One method of overcoming these difficulties is shown in Figure 4. A dual op amp is used to convert sense current to a negative output voltage, and then invert the negative voltage to produce a positive output. The negative power supply voltage is supplied by a simple and inexpensive 555 charge pump. Running at 60 kHz, this circuit converts the gate drive supply to a local negative rail that is capable of sinking roughly 50 milliamps. For this circuit, sensing equations 4, 5 and 6 apply with the negative signs removed from equations 4 and 5.

Accuracy

The inherent accuracy that is associated with splitting current between matched cells in a power MOSFET is relatively good. Assuming that both source and mirror terminals are held at the same potential, accuracy is solely

dependent upon the current mirror ratio, n. This parameter typically runs within ±1% of nominal at 25°C, is spec'd at ±3%, and remains within a ±3% window over temperature.

When SENSEFET devices are used in virtual ground sensing circuits, accuracy is a relatively straightforward issue. Current mirror tolerance adds with sense resistor tolerance and op amp offsets to produce a sense voltage that is easily maintained within ±5% over temperature.

Life gets much more interesting when a conventional mirror to ground sense resistor is used. Instead of deriving measurement accuracy from matched on–resistances within a monolithic device, tolerance depends both on internal ratios and the ratio of internal on–resistance to an external R_{SENSE}. Therefore, in this configuration, unit–to–unit variations and temperature stability are first order design considerations.

Referring again to Figure 2, the sensing model provides a schematic illustration of the issues involved. To start, let's assume that R_{SENSE} is equal to zero. In this condition, whatever variations occur in $R_{a(on)}$ are very nearly matched on a percentage basis by variations in $R_{DM(on)}$. Therefore, even for very large changes in $R_{a(on)}$ the ratio between $R_{a(on)}$ and $R_{DM(on)}$ remains nearly constant. Since this ratio is undisturbed, the ratio of sense current to drain current is also undisturbed, and measurement accuracy is relatively good. At the other end of the spectrum, let's assume that R_{SENSE} is an open circuit. In this case mirror voltage is not dependent upon ratios. The mirror terminal samples the voltage drop across $R_{a(on)}$, with $V_{SENSE} = I_D \times R_{a(on)}$. Measurement accuracy is, therefore, directly dependent upon the value of $R_{a(on)}$. Since $R_{a(on)}$ can vary 30% from unit to unit and 100% over temperature, an accurate measurement is not obtained in this configuration. In
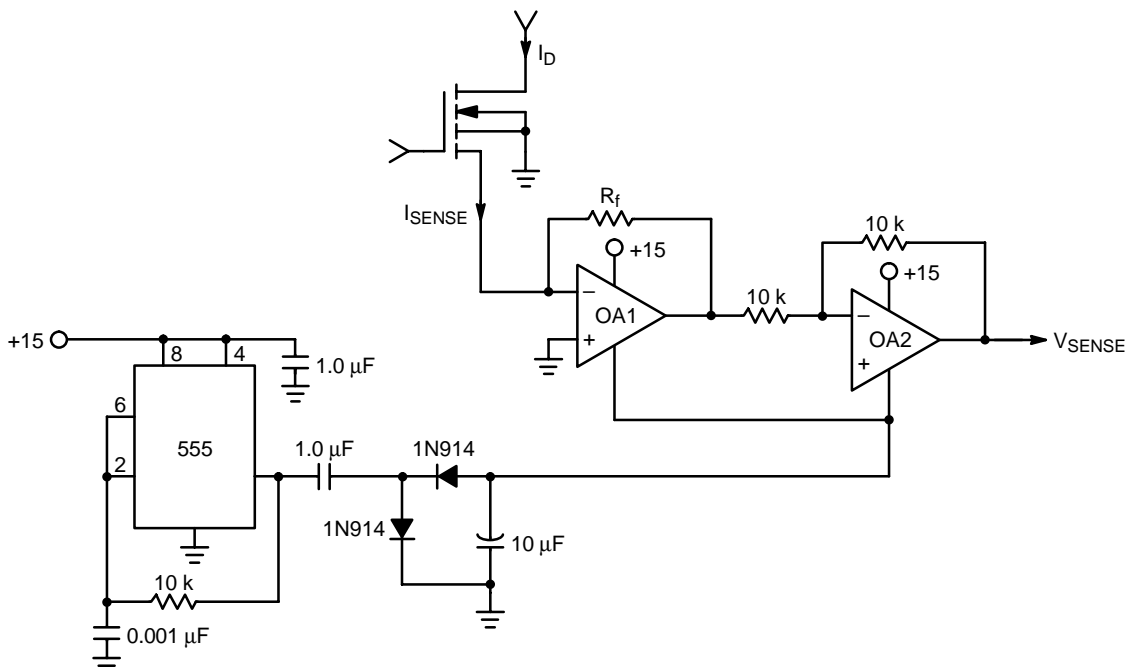

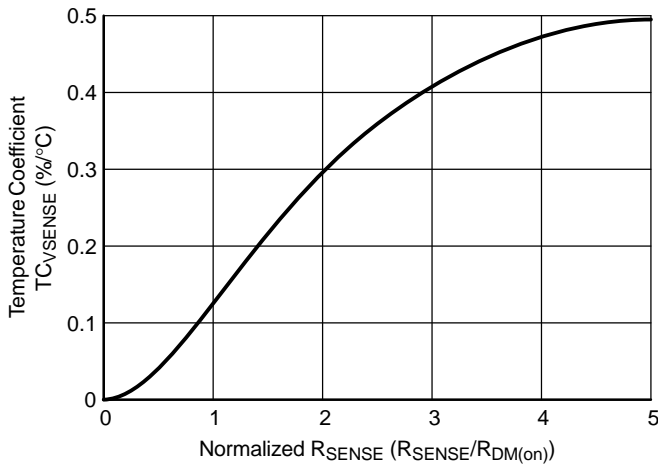**Figure 4. Positive Output Virtual Ground Sensing Circuit**
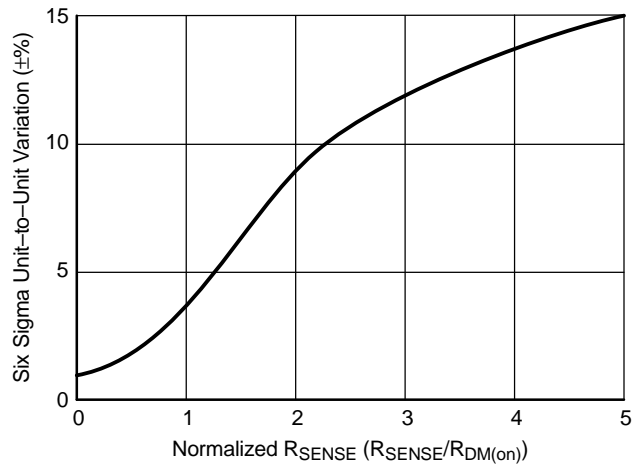
Figure 5. Temperature Stability

Figure 6. Unit–to–Unit Variation

between these two extremes, choosing R_{SENSE} boils down to a tradeoff between signal level and accuracy. As a rule of thumb, useful performance is obtained with values of R_{SENSE} up to $R_{DM(on)}$. Above $R_{DM(on)}$, however, measurement accuracy is more dependent upon the absolute value of $R_{a(on)}$ than it is on the ratio, and performance degrades rapidly as R_{SENSE} is increased further. An illustration is provided in Figure 5 where temperature stability is plotted versus normalized R_{SENSE}. In this figure, R_{SENSE} is normalized with respect to $R_{DM(on)}$ such that a value of 1 corresponds to $R_{SENSE} = R_{DM(on)}$. Note that temperature stability is relatively well behaved at values of R_{SENSE} that are below $R_{DM(on)}$ and rapidly degrades at higher values.

A similar curve that shows unit–to–unit variations versus R_{SENSE} is plotted in Figure 6. Again, the story is very much the same. At values of R_{SENSE} below $R_{DM(on)}$, performance is well within the kind of tolerances that are usually needed for current limiting circuits. Above $R_{DM(on)}$, however, unit–to–unit stability also rapidly degrades.
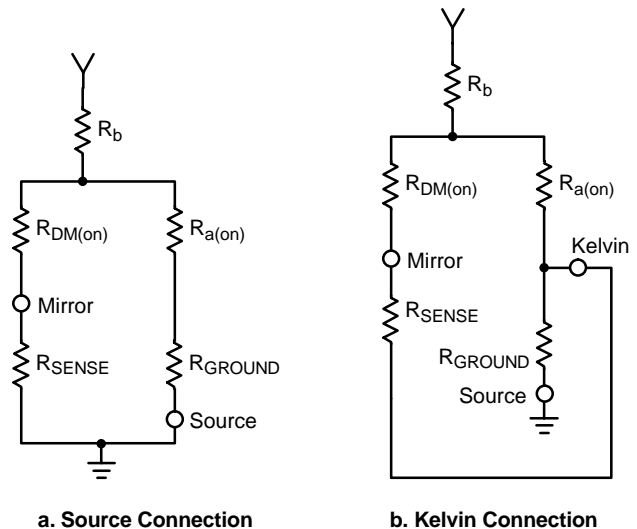
Kelvin Source Connection

In order to get the full accuracy that SENSEFET devices are capable of, a Kelvin connection to the source is required. Otherwise voltage drops that are caused by load current flowing in the ground connection will add to sense voltage and introduce a source of error. The effect of ground impedance is illustrated in Figure 7 where R_{GROUND} has been added to the model in Figure 2. Load current flowing through R_{GROUND} produces a voltage drop that appears in series with $R_{a(on)}$. In Figure 7a, this voltage adds directly to the open circuit mirror voltage, and affects the measurement proportionately. In Figure 7b, the Kelvin connection removes parasitic ground voltage from the measurement by referencing R_{SENSE} directly to the devices source metallization.

The impact upon measurement accuracy is largely a function of the device ratings and circuit layout. Devices such as the MTP10N10M and MTP10N25M that have

$R_{a(on)}$'s in excess of 100 milliohms are affected less than 10% by the roughly 10 milliohms of parasitic resistance that occurs from the source metallization of a TO–220 device to a soldered PC board connection. Therefore, with these devices a Kelvin return makes only a minor difference. The same 10 milliohms of parasitic resistance however, introduces quite a substantial error for higher current devices such as the MTP40N06M. For this device, 10 milliohms of parasitic resistance is an appreciable fraction of its 17 milliohm $R_{a(on)}$, and use of a Kelvin return eliminates what would otherwise be a large error.

Although SENSEFET Kelvin returns are intended primarily to improve measurement accuracy, they also have an important application at higher frequencies. As illustrated in Figure 8, gate drive ground can be referenced to the Kelvin pin instead of a PC board ground plane. This connection bypasses parasitic source inductance, L_S, that is associated with the source wire bond and the source pin. Therefore, the Kelvin connection allows faster switching speeds at higher currents where source inductance becomes a significant limitation.

a. Source Connection   b. Kelvin Connection
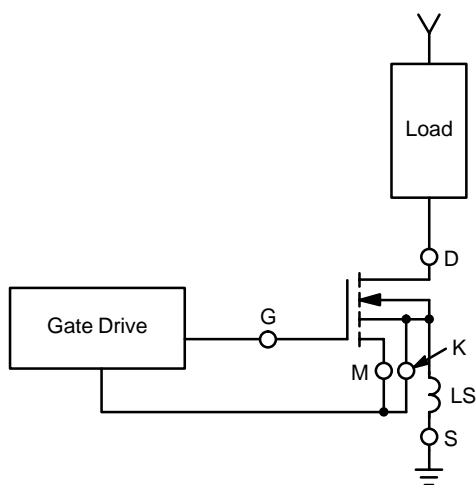
Figure 7. Ground Impedance

Figure 8. High Speed Gate Drive

To illustrate, let's assume that a conventional TO–220 MOSFET is connected such that the only source inductance in its gate drive loop is 8 nanohenries of wire bond and lead inductance. During switching transitions, the voltage drop across this inductance ($V = L_S \, dI_S/dt$) opposes gate drive voltage. As long as the magnitude of this voltage is less than the amount by which the gate is overdriven, no first order effect upon switching times is seen. However, when this voltage becomes high enough to significantly oppose the gate drive, there is a limiting effect upon switching speed.

Since the speed at which a MOSFET's gate can be driven is essentially independent of drain current, switching speed tends to be relatively independent of drain current, and dI_S/dt increases with increasing drain current. However, at a critical value of drain current $L_S(dI_S/dt)$ becomes large enough to appreciably interfere with gate bias. At this point, a significant slowing of switching occurs as the parasitic opposing voltage debiases the gate by the amount of gate overdrive voltage. For TO–220 MOSFETs, this limitation becomes significant between 10 and 20 amps, where transition times that can easily be held below 10 ns at 10 amps climb to several tens of ns at 20 amps and above.

Table 2. High Speed Performance (Fall Time)

High Speed Performance		
Drain Current Amps	Kelvin Return Fall Time (ns)	Source Return Fall Time (ns)
5	6	6
10	6	7
15	6	9
20	7	27
25	9	28
30	10	44

If a SENSEFET device is substituted for a conventional MOSFET and the KELVIN gate drive return of Figure 8 is used, then the parasitic voltage occurs outside the gate drive loop and has substantially less effect. The difference is illustrated in Table 2 where switching times for an MTP10N25M with Kelvin gate drive return and Source gate drive return are compared. In both cases, the test device is switching an inductive load, is driven by a high speed MC33152 driver and is measured with a Pearson model 4028 high speed current transformer.

It is readily observed from this data that SENSEFET packaging contributes a significant performance advantage at higher currents, but has no appreciable effect at currents below 10 amps.

Noise Suppression

Particularly at high speeds, noise spikes at both turn–on and turn–off are a first order design consideration in SENSEFET circuits. These spikes are short, roughly the same duration as the switching transitions that produce them, and can be several times the value of the sense voltage that is being measured. They arise from current flow in parasitic capacitances and also from higher sensing gain as the device switches through its active region.

Generally speaking, noise at turn–off is not usually much of a problem, since a spike that is large enough to cause current limiting has no effect during turn–off. Turn–on, however, is another matter. Spikes that are routinely seen at turn–on can easily produce false trips that interfere with proper circuit operation. Fortunately, filtering is relatively easy. Due to the limited duration of the noise, a simple single pole RC filter is quite effective. Using an op amp instead of a comparator to monitor sense voltage also works well. The op amp's roll off characteristic provides a single pole filter that causes short spikes to be ignored. Both of these techniques, of course, slow the speed of the current limit loop and may be undesirable in some applications. In these instances digital blanking can be used to disable the current limit comparator for a fixed amount of time during the turn–on switching interval.

Whatever circuit technique is used to limit noise, layout is critical when SENSEFET devices are used at or near their fastest switching speeds. Printed circuit board construction is a must, and careful attention is required to keep ground currents away from the current sensing loop. Small ground planes, separate grounds for power and sense loops, and single point grounding for power components all tend to make life easier. It is also important to remember that radiated noise can be as big a problem as conducted noise. For this reason it is important to minimize the size of the power ground, which is a radiating surface, and to place sense voltage filters as close as possible to the current limit circuits that they feed.

Sense Voltage Under Abnormal Conditions

For normal steady–state operation, sense voltage is determined by resistor ratios which are easily calculated. However, questions often arise concerning sense voltage under abnormal conditions. Operation in avalanche, operation at very low currents, reverse current flow, and reverse recovery of the drain–source diode are considered here.

Avalanche

The latest generation of power MOSFETs is capable of withstanding considerable stress in avalanche. Drain–source diodes have been replaced with drain–source Zener transient suppressors. Therefore, inductive fly–back voltages can be clamped directly by the MOSFET. SENSEFET product can also be used in this way, which leads to questions regarding sense voltage during avalanche.

The simplified model in Figure 9 provides a basis for analysis. It assumes one sense cell, n power cells, and a bulk drain resistance which is negligible. In avalanche individual cells look like paralleled Zeners, each of which has a small series impedance R_Z. Although R_Z is probably well matched from cell to cell, it is well known that epi thickness, and therefore breakdown voltage, vary across the surface of a semiconductor wafer. From point–to–point on a power transistor chip, the variation can exceed a full percent. Therefore, the Zener voltages in Figure 9 can be mismatched by a volt or more in situations where the average voltage across R_Z is measured in tens or hundreds of millivolts. Under these conditions, sense voltage is essentially indeterminate since I_{SENSE} will depend more upon the sense cell's relative breakdown voltage than on the current that is being measured. The net result is that energy rated SENSEFET product will survive avalanche as well as energy rated MOSFETs, but lose their ability to provide a meaningful current sensing measurement in this condition.
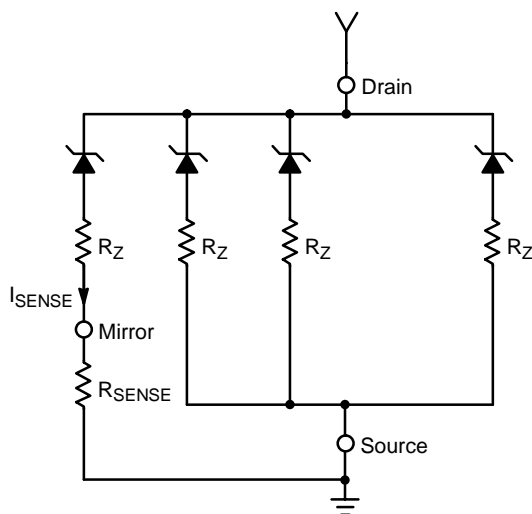

Figure 9. Avalanche Model

Very Low Drain Current

At normal operating currents, offset voltage in measuring circuitry is usually negligible. However, at drain currents below 1% of rated output current, offset voltages have a first order effect on the current sensing measurement. A brief look at the simplified model in Figure 10 explains why. Suppose, for example, that the device is an MTP10N25M with an $R_{a(on)}$ of 170 milliohms and an $R_{DM(on)}$ of 145 ohms. With one amp flowing through $R_{a(on)}$ and zero offset in the measurement amp, 170 millivolts is impressed across $R_{DM(on)}$ to develop I_{SENSE}. If the offset voltage of the measurement amp is 1 millivolt, the voltage across $R_{DM(on)}$ is changed by 1 millivolt. Sense current is changed by 1 part in 170, which is less than 1%. Now assume that I_D is 10 milliamps. In this case, only 1.7 millivolts is developed across $R_{a(on)}$. Therefore, a 1 millivolt offset changes the measurement by 1 part in 1.7, and seriously affects the results. In general, at very low currents measurement amplifier input offset characteristics significantly influence measured current mirror ratios, and can cause relatively large deviations from values observed at practical operating currents.
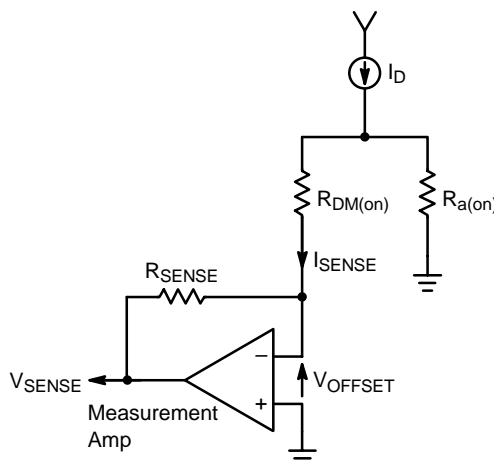

Figure 10. Low Drain Current

Reverse Recovery

When the devices are used in bridges, current flow in the drain–source diode can be an important part of circuit operation. If this diode is commutated, in other words switched off rapidly by an opposite half bridge, a significant amount of sense voltage is produced. For example, suppose that inductive load current is freewheeling through a SENSEFET's drain–source diode as shown in Figure 11 and then S1 closes rapidly. As the accompanying waveforms suggest, a very substantial V_{SENSE} spike is produced as the drain–source diode is cleared. Its magnitude is dependent upon the speed at which S1 closes, the amount of current that is commutated, and the value of R_{SENSE}. Its duration is a function of the drain–source diode's reverse recovery time.
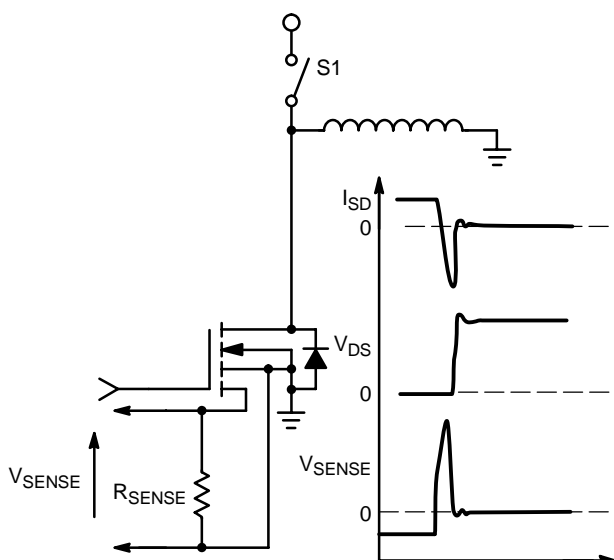
Figure 11. Reverse Recovery

Figure 12. Reverse Current Flow

This spike is a significant design issue because it is both large and occurs as a positive signal with respect to ground. To illustrate the magnitude of the problem, consider an MTP10N25M in which 2 amps is commutated in 100 ns. With R_{SENSE} = 100 ohms, the reverse recovery spike has a magnitude of 400 millivolts. Since the forward sense voltage at 2 amps is only 140 millivolts, the possibility of falsely tripping a current limit circuit at very modest currents exists.

Fortunately the duration of the spike is very short. It is limited to the time during which current is being commutated in the drain–source diode. Therefore, a single pole RC filter will easily reduce peak voltage to acceptable levels. Alternately, current limit circuitry can be digitally blanked during the time that reverse recovery transitions take place.

Reverse Current

When current flows through the drain–source diode, anode current is spilt between source and mirror terminals in a fashion similar to forward mode operation. However, the equivalent circuit for reverse operation is significantly different, as are operating results. As shown in Figure 12, current sensing is dependent upon mirror–to–bulk and source–to–bulk diodes as well as the ratio of cell resistances. As long as R_{SENSE} is equal to zero or effectively help to zero with an op amp, the sense ratio is based upon the ratio of R_{mb} to R_{sb} and matched diode drops. Since the ratio of R_{mb} to R_{sb} is the same as the ratio of $R_{DM(on)}$ to $R_{a(on)}$, forward and reverse sense ratios match and are equal to n, the current mirror ratio.

When a non–zero R_{SENSE} is added, a significant departure from forward mode operation is observed. Additional resistance in the sense leg does not give the same linear predictable results, because cell resistances in the reverse mode are modulated by minority carrier current flow. These resistances have a first order dependence upon current density and decrease in value at higher currents.
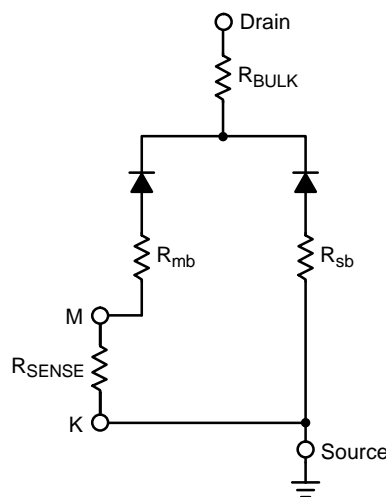
Therefore, while matching from cell–to–cell with R_{SENSE} = 0 is quite good, any attempt to establish a ratio between the internal resistances and a fixed external R_{SENSE} is met with first order non–linearities. The net result is a measurement that is accurate within a few percent if a virtual ground op amp is used, and is a poor representation of reverse current flow when practical values of sense resistance are connected from mirror to Kelvin.

SENSEFET Compatible Integrated Circuits

Since power MOSFETs are driven by integrated circuits in many applications, compatibility with integrated circuits is an important issue. As a general rule, current limit comparators with 100 to 125 millivolt sensitivity are required for PWM integrated circuits to be compatible with SENSEFET devices. A number of recently developed integrated circuits either meet this requirement or were developed specifically to work with this product. They are listed as follows:

Device	Description	Application
MC33152	SENSEFET Driver	Microprocessor Interface
MC33034	Motor Controller	Brushless DC Motors
MC34129	Current Mode Controller	Single Transistor SMPS
SG3526	Power Supply Controller	Push–Pull SMPS

All of these devices function quite comfortably with output voltages that the devices provide. They are also all capable of driving SENSEFET product at very high switching speeds. Therefore, good PC board layout techniques are very important when using these circuits. In addition, the use of series gate resistors to slow switching speed can be very helpful for breadboarding and debugging.

Two examples which illustrate how SENSEFET product is used appear in Figures 13 and 14. Figure 13 describes an isolated 12 volt to 5 volt current–mode supply, and is a convenient vehicle for showing how the MC34129 and MTP10N10M work together.
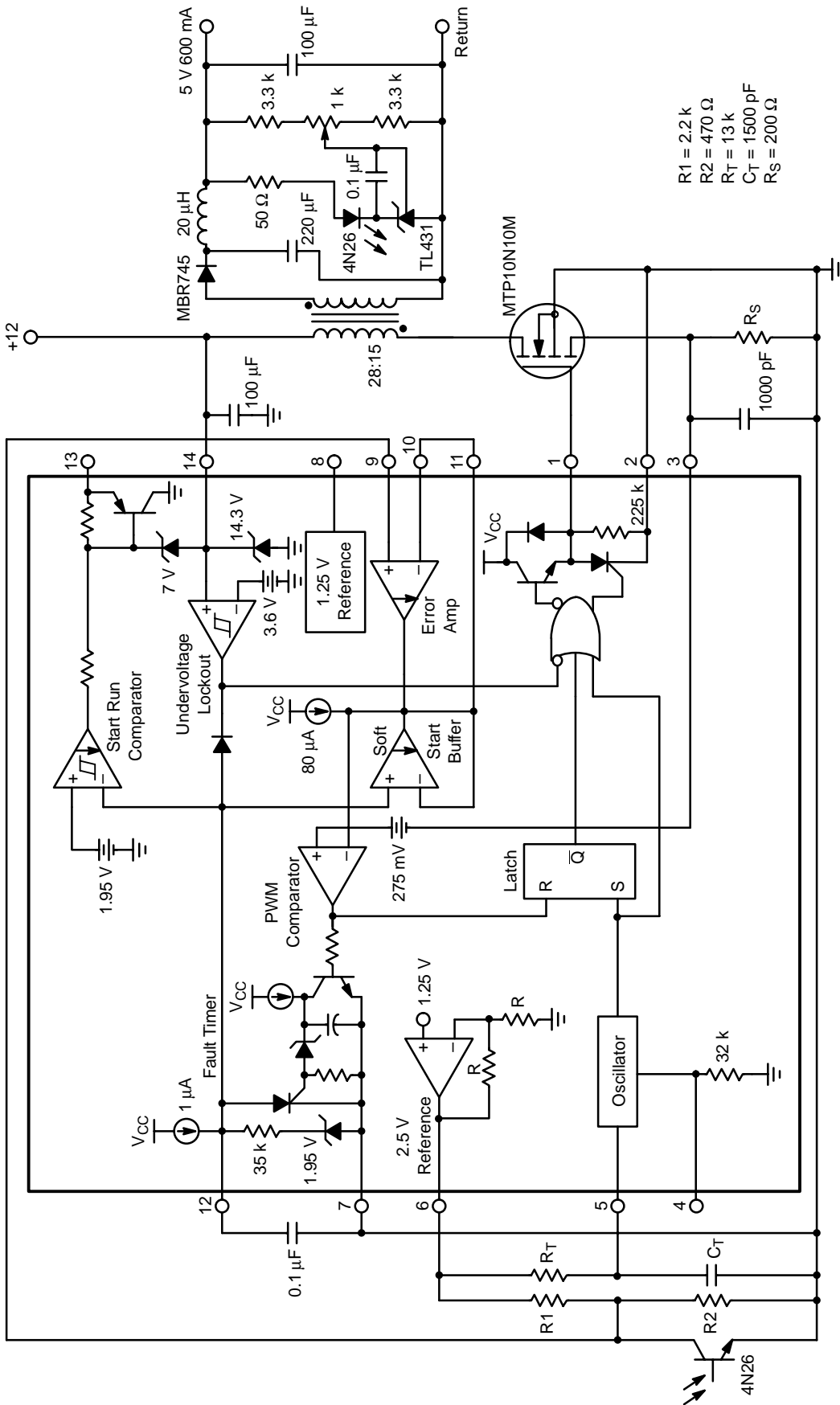
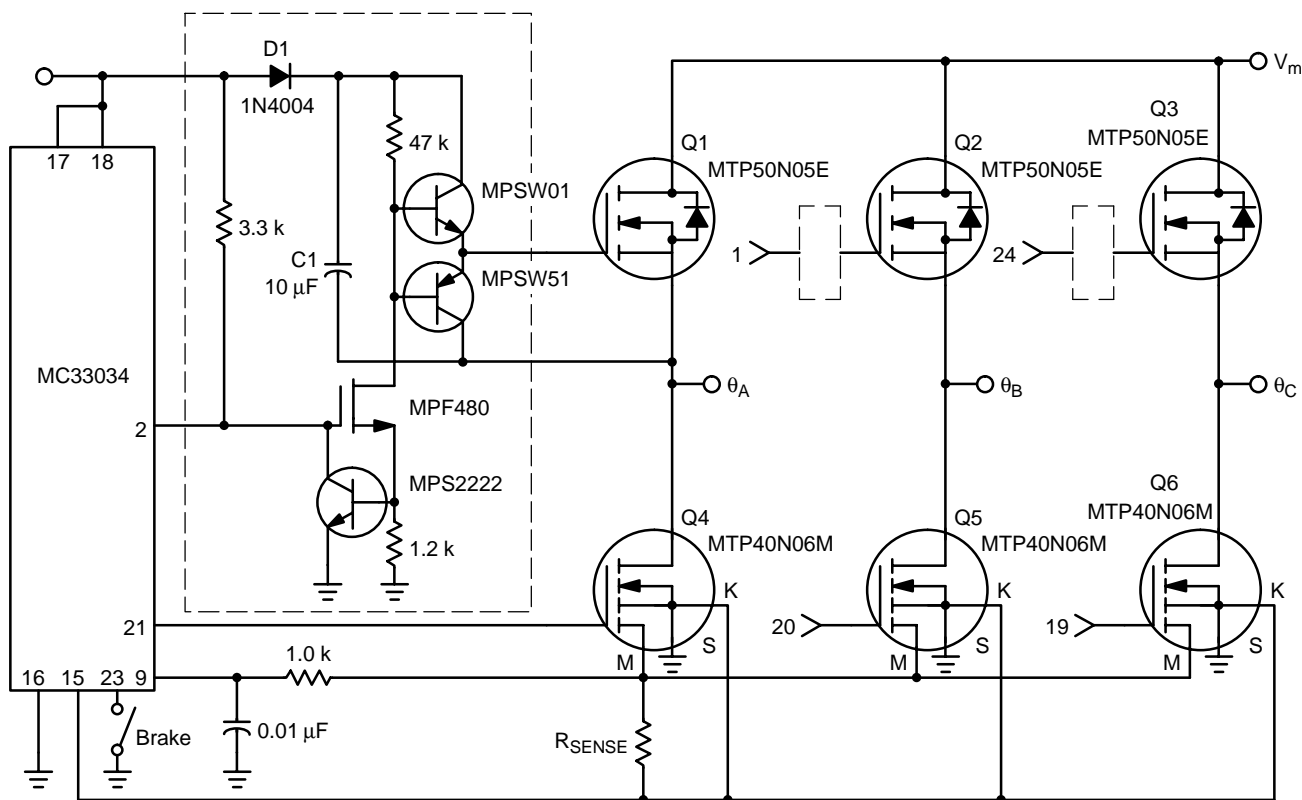**Figure 13. MC34129 Application Example**

Figure 14. MC33034 Application Example

Starting with the oscillator, R_T and C_T are selected for operating frequency and dead time. A combination of 13 k for R_T and 1500 pC for C_T produces 28 kHz operation with slightly less than 50% maximum duty cycle. Ramp voltage is generated by R_S and fed into the PWM comparator's non–inverting input at Pin 3. The ramp's magnitude is determined by using equation 1 and adding –120 μA of nominal input bias current that flows out from Pin 3. The ramp voltage is, therefore:

$$V_{RAMP} = 57 \text{ mV/Amp} \cdot I_D + 24 \text{ mV}$$

Knowing the relationship between V_{SENSE} and primary current, maximum short circuit current is set with voltage divider R1/R2. The output voltage from this divider is coupled through a unity gain Error Amp to set the PWM comparator's upper trip point. To calculate the trip point, 275 mV of offset is added to the output voltage of the SENSEFET device. With the values shown, R1 and R2 set the upper trip point at 470 mV, and peak current is limited to approximately 2.8 amps.

Turning to the brushless motor drive in Figure 14, an illustration of how the MC33034 brushless motor controller and MTP40N06M interface with each other is provided. In this figure, six power MOSFETs are connected in a 3–phase bridge configuration to drive a brushless motor. The upper devices are 28 milliohm N–Channel FETs that are driven by a bootstrap technique that is described in detail in AR194. To form the lower half bridge, three MTP40N06M are directly driven by the MC33034.

Since only one of these devices is on at a given time, all three mirror terminals are connected to one sense resistor, R_{SENSE}. With this arrangement, a current–trip threshold is reached when excess current appears in any of the three phases. A single pole RC filter is inserted between R_{SENSE} and the MC33034's current limit comparator to eliminate the noise spikes that inevitably occur at R_{SENSE}.

The MC33034's current limit threshold is 100 mV. Therefore, from equation 2, the value of R_{SENSE} that is required for a desired current limit can be calculated as following:

$$R_{SENSE} = 0.1 \text{ V} \cdot R_{DM(on)}/[(I_{limit} \cdot R_{a(on)}) - 0.1 \text{ V}]$$

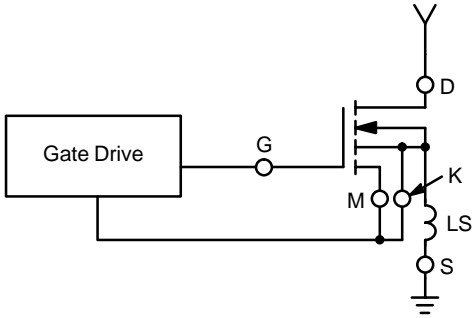Plugging in $R_{a(on)}$ and $R_{DM(on)}$ for the MTP40N06M:

$$R_{SENSE} = 0.1 \text{ V} \cdot 16 \text{ ohms}/$$
$$[(I_{limit} \cdot 0.017 \text{ ohms}) - 0.1 \text{ V}]$$

To obtain a current limit value of 40 amps, for example, these values lead to choosing a value of 2.7 ohms for R_{SENSE}. Since this resistor can be a standard 1/4 watt resistor, it is by far preferable to the alternative of using a 2.5 milliohm 4 watt resistor in series with the ground return.

SENSEFET PRODUCT FOR HIGH FREQUENCY APPLICATIONS

Although current sensing power MOSFETs are intended primarily for lossless current sensing, they also have attractive characteristics for high frequency applications. In particular, packaging is superior for high frequency

circuits since a Kelvin connection is provided to the device's source metallization. As illustrated in the accompanying figure, gate drive ground can be referenced to the Kelvin pin instead of a PC board ground plane. This connection bypasses parasitic source inductance, L_S, that is associated with the source wire bond and the source pin. Therefore, the Kelvin connection allows faster switching speeds at higher currents where source inductance becomes a significant limitation.

To illustrate, let's assume that a conventional TO–220 MOSFET is connected such that the only source inductance in its gate drive loop is 8 nanohenries of wire bond and lead inductance. During switching transitions, the voltage drop across this inductance ($V = L_S \cdot dI_S/dt$) opposes gate drive voltage. As long as the magnitude of this voltage is less than the amount by which the gate is overdriven, no first order effect upon switching times is seen. However, when this voltage becomes high enough to significantly oppose the gate drive, there is a limiting effect upon switching speed. For TO–220 MOSFETs, this limitation becomes significant between 10 and 20 amps, where transition times that can easily be held below 10 ns at 10 amps climb to several tens of ns at 20 amps and above.

If a SENSEFET device is substituted for the conventional MOSFET and a Kelvin gate drive return is used, then the parasitic voltage ($V = L_S \cdot dI_S/dt$) occurs outside the gate drive loop and has substantially less effect. The difference is illustrated in the following table, where

switching times for an MTP10N25M with Kelvin gate drive return and Source gate drive return are compared. In both cases, the test device is switching an inductive load, is driven by a high speed MC33152 driver, and is measured with a Pearson model 4028 high speed current transformer.

High Speed Performance		
Drain Current Amps	Kelvin Return Fall Time (ns)	Source Return Fall Time (ns)
5	6	6
10	6	7
15	6	9
20	7	27
25	9	28
30	10	44

It is readily observed from this data that SENSEFET packaging contributes a significant performance advantage at higher currents, but has no appreciable effect at currents below 10 amps.

Conclusion

SENSEFET products are conceptually simple devices that provide an alternative to power sense resistors and magnetics for sensing load current. When suing these devices, calculating sense resistance and sense voltage for steady–state conditions is relatively straightforward. Other aspects of designing with this technology requires greater familiarity with both characteristics and techniques for dealing with relatively low sense voltages. One subtle characteristic provides an unintended benefit. The Kelvin source connection can be used to attain faster switching speeds than similar devices in conventional three leaded packages. Whether they are used for speed or for lossless current sensing, current sensing power MOSFETs are well worth becoming familiar with; since circuits based upon SENSEFET designs tend to operate more efficiently with fewer and smaller components.

Notes

SENSEFET is a trademark of Semiconductor Components Industries, LLC (SCILLC).

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