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By: Kenneth Dierberger
Richard Redl
Leo Saro

APPLICATION NOTE

High-Voltage MOSFET Behavior in Soft-Switching Converters: Analysis and Reliability Improvements

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High-Voltage MOSFET Behavior in Soft-Switching Converters: Analysis and Reliability Improvements

Leo Saro
SICON S.R.L.
V. della tecnica, 1
36030 Villaverla (VI)
Italy
pad@sicon-ups.com

Kenneth Dierberger
Advanced Power Technology
405 S. W. Columbia St.
Bend, OR 97702
U.S.A.
KenD@advancedpower.com

Richard Redl
ELFI S.A.
En Montévaux
CH-1726 Farvagny-le-petit
Switzerland
RichardRedl@compuserve.com

Abstract -The benefits of soft-switching converters, and in particular zero-voltage switching (ZVS, also called zero-voltage-transition [ZVT] or resonant-transition) circuits, are well-known. High-frequency converters powered from high source voltage show significant improvements when operated with soft switching. These improvements are 1) reduced switching losses, which allow high switching frequency and size reduction of reactive components, 2) reduced EMI/RFI noise, 3) no need for complex and expensive snubbers, and 4) exploitation of the parasitic circuit elements to help the resonant transition. Thanks to these characteristics, zero-voltage-switching topologies are now widely used in power electronics, and especially in telecom power systems.

The MOSFET is the most common choice of controlled switch in the zero-voltage-switching full-bridge converter. The MOSFET is capable of very fast commutations and its intrinsic body diode saves an additional external component that would otherwise be necessary to clamp the switch voltage to the input supply voltage. Both the internal body diode and the output capacitance become essential components for the resonant transition.

Although in the zero-voltage-switching full-bridge converter the MOSFET operates well inside its safe operating area and its body diode is never subjected to hard turn-off, some "unexplainable" failures do happen, due to the unavoidable usage of the intrinsic body-diode.

This paper analyses the MOSFET's behavior in high-power and high-input-voltage ZVS converters, and presents an original theory of the MOSFET breakdown.

New technical solutions to improve the ruggedness of the device and consequently the reliability of the equipment are proposed. The effectiveness of these solutions is proved in a 100-A, 6000-W telecom power supply.

1. Introduction

A 60-V, 100-A rectifier for telecom applications was developed using 1000V power MOSFETs in the ZVS dc/dc converter section. Originally standard recovery devices were employed based on the valid assumption that the intrinsic body diodes have never been exposed to hard turn-off, that is the body diodes were never reverse-polarized while current was flowing in them.

During mass production we experienced a number of MOSFET failures in the ZVS converter. The subsequent investigation definitely proved that the device operated well inside its SOA. In fact, we could demonstrate that under all conceivable normal and abnormal conditions all the MOSFET static and transient electrical parameters stayed within their absolute maximum ratings.

The failure analysis, performed on a statistically significant population, provided the following information:

1) The lower the body-diode reverse recovery time is, the lower the failure rate will be. Note that we could verify, by installing a detecting and recording circuitry in all the power supplies, that the MOSFET body diode was never subjected to hard turn-off.

2) All the failures happen when the output load is less than 25% of its maximum value, in spite the fact that in this case the MOSFETs are less stressed (i.e., they run with lower current, power dissipation, dv/dt, overvoltage, etc.) than when the load is higher.

3) Equipment that were working below the critical light load for a long enough time, have a good chance to continue working without a MOSFET failure.

4) Normally, if an equipment had a MOSFET failure, it works without any relevant failure after the replacement of the failed device.

This paper attempts to explain the above information; it also presents solutions to eliminate the MOSFET failures.

Section 2 of the paper presents a short description of the operation of the ZVS full bridge converter, highlighting the body-diode behavior.

Section 3 describes the MOSFET body diode.

Section 4 explains the MOSFET breakdown theory and how it accounts for the facts listed above.

Section 5 provides results of experiments performed on the body diode with the purpose of verifying the breakdown theory.

Section 6 describes new technical solutions to improve the reliability of the ZVS converter and presents experimental results.

2. Basic Operation of the PWM Phase-Shift Zero-Voltage-Switching Full-Bridge Converter

The basic circuit (Fig. 1, [1], [2], [3], [4]) comprises four switches labeled S1 through S4, each shown shunted by itsbody diode and output capacitance. An additional small inductance and two small clamp diodes [1] are employed in the primary side of the transformer to help maintaining soft switching at reduced load and limiting the voltage overshoot of the output rectifier diodes.

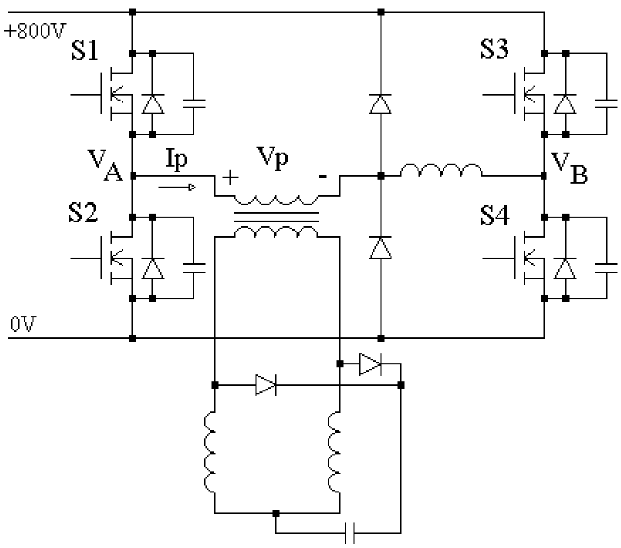


Fig. 1. Zero-voltage-switching full bridge converter.

The converter operates with a fixed switching frequency of 50 kHz and with a regulated 800-V supply voltage. The output rectifier is a current doubler. The effective duty ratio (defined for the voltage across the transformer) is determined by the phase between the commutation of the two legs of the bridge. In this converter, rather than modulating the actual pulse width of the gate drivers, we maintain the gate (and switch) duty ratios at 50% and vary the commutation time of the S3-S4 leg.

The switches of the converter operate with **lossless zero-voltage turn-on** and with **reduced-loss turn-off.** Due to the inductive load of a bridge leg and the current flowing in the load, there will be a natural commutation of the current from a switch to the body diode of the other switch in the same leg, leading to virtually zero voltage across the switch prior to turn-on. This eliminates the power losses caused by (1) the simultaneous presence of large current and voltage in a switch at each transition and (2) the discharge of the output capacitance of the switch. The output capacitance acts as a capacitive snubber, reducing the turn-off loss caused by the finite turn-off time of the switch. The parasitic circuit elements (output capacitance, leakage and magnetizing inductance, body diode) are used advantageously to facilitate low-loss resonant-transition switching.

Switches in this converter are employed differently than those of a hard-switching standard converter, since the MOSFET's internal body diode and output capacitance become essential components in the resonant transition.

The timing diagram (Fig. 2) shows the operation of the converter. There are seven states in a full switching cycle. The states are as follows:

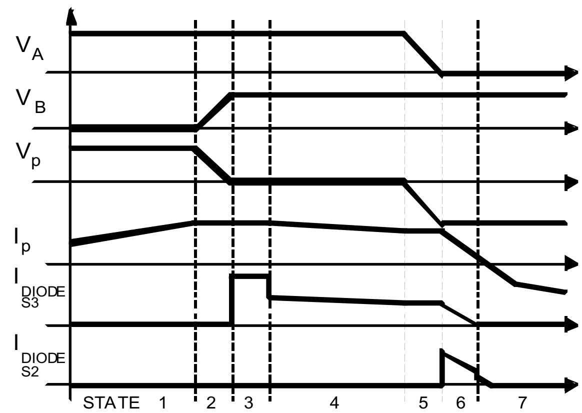


Fig. 2. Timing diagram of the full-bridge zero-voltage-switching converter.

- State 1) Powering

The switches **S1 and S4 in the diagonal are on**. V_A = 800 V, as S1 is on, V_B = 0 V, as S4 is on, V_P is equal to the entire DC voltage, 800 V.

During this stage the primary current I_P of the transformer increases because of the contribution of magnetizing current and especially of the current in the output inductors. The body diodes of the switches are not conducting. The power is transferred from the input to output. This is the **ON** portion of the cycle.

- State 2) Resonant transition: Powering to Free-wheeling

At the end of the powering state, the switch **S4 is turned off**. Since the inductor current cannot change abruptly, it continues to flow, changing the voltages of the output capacitances of S3 and S4, and so raising V_B towards the DC rail.

- State 3) Free-wheeling (body diode)

When V_B reaches the DC voltage, **the body diode of S3 begins to conduct**, clamping V_B to 800 V. The total primary current I_P is now flowing through the body diode.

- State 4) Free-wheeling (body diode + channel)

A few hundred nanoseconds after the body-diode of S3 has begun to conduct, **S3 is turned on**, in order to lower the conducting loss by placing the MOSFET's conducting channel in parallel with the body diode. Note that the turn-on of S3 takes place when the voltage across it is zero, therefore the turn-on becomes lossless.

A portion of the (negative) primary current I_p is now flowing in the channel of S3; the remaining current flows in the body diode. All the current of S1 is positive and flows in its channel. V_A = 800 V, as S1 is on, V_B = 800 V, as S3 is on, and V_P = 0 V.

This is the **free-wheeling** portion of the cycle. It allows the fixed frequency operation; in fact, this situation remains until the beginning of the other diagonal powering step (S2 and S3 conducting). This portion normally lasts for a couple of microseconds.

- State 5) Resonant transition: Free-wheeling to Powering

S1 is turned off and its current is diverted from the channel to the output capacitance. If the channel current falls to zero before V_{DS} has substantially risen, we obtain lossless turn-off commutation. The current now changes the voltages of the output capacitances of S1 and S2, driving V_A from 800 V towards 0 V.

- State 6) Resonant transition: Free-wheeling to Powering (body-diode)

When V_A reaches zero voltage, **the body diode of S2 begins to conduct**, clamping V_A to 0 V. The whole primary current I_P is now flowing in the body diode of S2 and in the body diode and channel of S3. V_A = 0 V, as S2 is on, V_B = 800 V, as S3 is on, and V_P = -800 V. The current is quickly decreasing as the transformer and external inductors are now reverse polarized.

- State 7) Powering

State 6) lasts for a few hundred nanoseconds, until **S2 is turned on**. Once again turn-on takes place when the voltage across the device is zero, obtaining lossless commutation.

The current through both the S2 and S3 body diodes falls quickly to zero, because the MOSFET channel is diverting part of the current, and especially because the transformer is polarized to reverse the direction of the current I_p.

When the inductors have completed their charge, power is transferred to the output and the converter is in a situation like step 1). This is another ON state. With the parameter values of our design, it lasts for about 7 μsec. Now the converter is ready to repeat another cycle, very similar to this described above, this time with the S2 - S3 diagonal conducting.

As we can see, the MOSFET is always turned off only after the current in its body-diode has reversed and has been flowing directly in the MOSFET channel for several microseconds; consequently the application of high dv/dt to the body diode happens several microseconds after it stopped conducting.

3. Description of the Body Diode

All power MOSFETs have an intrinsic bipolar transistor in their structure. The vertical DMOS device, as illustrated in Fig. 3, has the base-emitter junction of the bipolar transistor shorted by the source metallization, forming the "Body Diode".

If this parasitic bipolar transistor becomes active, the classic mechanism of second breakdown with current hogging can occur. This is well known in literature and reported also in MOSFET application notes [5], [6]. The

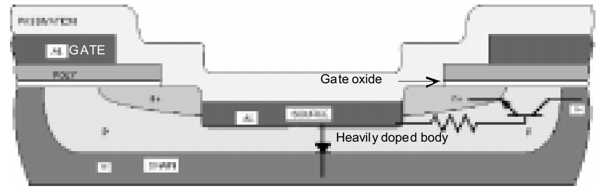


Fig. 3. Vertical DMOS cross-section

current hogging causes local heating, thereby increasing the bipolar gain, further constricting the current, and eventually leading to device failure. If the body diode conducts in the forward direction, minority carriers remaining in the base region during diode recovery can cause transistor action with destructive results.

4. Failure Mechanism Theory

It seems to be a common belief that a zero-voltage-switching (ZVS) topology, where the MOSFET is turned on while the body diode is conducting, the body diode will not be subject to second breakdown as current is reversed for a period of time during the cycle which should be long enough for the diode to recover.

The reality is that charge will remain stored in the body diode for a period of time much longer than the data sheet value for reverse recovery time or until high voltage is applied which will sweep the minority carriers out of the junction. Therefore, when reverse high voltage is applied to the body diode second breakdown may still occur even after a relatively long time has passed.

In a typical ZVS topology, forward current is forced into the body diode to clamp the output to either the positive or negative rail. This forward current causes the generation of minority carriers in both the P-body and N-epi regions (Fig. 4).

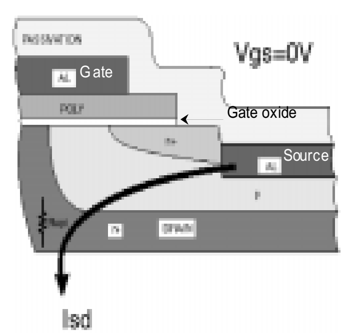


Fig. 4. Forward current flow in the body diode.

Next the MOSFET channel is turned on and diverts a portion of the current through the channel away from the body diode. A MOSFET channel can conduct current in both directions. The diversion of current away from the body diode will reduce the generation of minority carriers but will not stop it (Fig. 5).

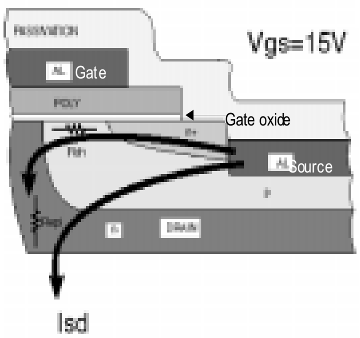


Fig. 5. Forward current flow in the body diode and the channel.

Next the external circuitry reverses the current flow through the device. This causes a small amount of reverse current flow in the body diode. The reverse current is small due to the very weak voltage field created by the low voltage generated by current flow in the low resistance of the channel. As a result some minority carriers will be swept across the junction removing them from the junction (Fig. 6).

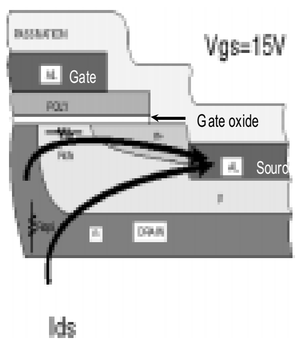


Fig. 6. Reverse current flow in the body diode and the channel.

After a short time enough carriers will have been removed enabling the junction to block a small amount of voltage. At this time most of the current is diverted to the resistive channel, developing a positive voltage across the $RDS_{(on)}$ that slightly reverse-biases the diode junction. Some carriers will continue to be eliminated by the normal recombination process and by a forced process that depends on the voltage applied across the diode (Fig. 7).

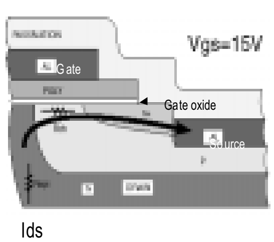


Fig. 7. Reverse current flow in the channel only.

The magnitude of this voltage depends on the primary current and consequently on the output load. If this voltage is small, a large number of minority carriers remain in the junction for a considerable time. When the channel is turned off, the MOSFET will begin to block voltage imposing a higher reverse voltage on the body diode. The application of high reverse voltage on the body diode will sweep the remaining carriers across the junction (Fig. 8). If the reverse current builds to a magnitude sufficient to activate the intrinsic bipolar transistor second breakdown may occur destroying the MOSFET.

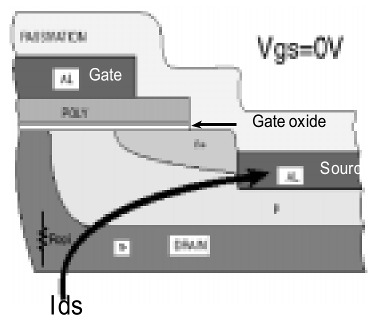


Fig. 8. Reverse current flow in the body diode only.

At light load (that is at low current in the MOSFET) the low positive voltage drop on the channel resistance may not be high enough to force the complete recombination of the minority charges before the end of the on time. This is especially true when the reverse recovery time of the body diode is long.

On the other hand, at heavy load (that is at high current in the MOSFET) the increased voltage drop on the channel resistance can be high enough to ensure a complete recombination of the minority charges during the on time, allowing a safe MOSFET turn-off at the end of this period.

Let us assume a statistical distribution, gaussian for instance, of the MOSFET population concerning the recombination time, its dependence on reverse voltage applied to the diode, the gain of the NPN parasitic device, its ruggedness to dv/dt, etc. It results that, under the same working conditions, only a percentage of the total devices used could be involved in our problem. These MOSFETs prone to fail must belong to the slower devices of all these statistical distributions. So we can also explain why:

∝ The faster the body-diode reverse recovery time, the lower the failure rate: *trr is a good indication of the minority charge recombination time, and this parameter is very important in our hypothesis. Qualitatively, we have the following situation (Fig. 9):*

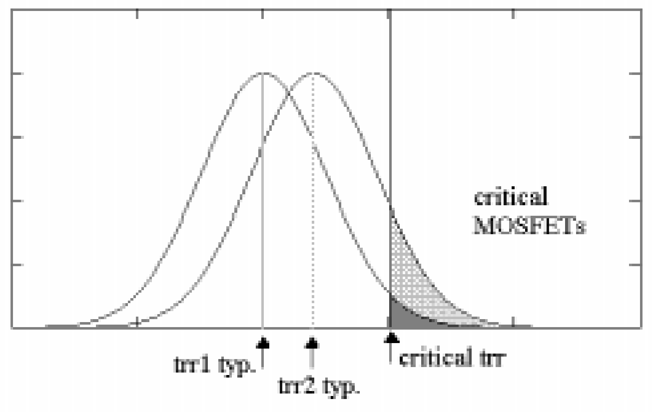


Fig. 9. Critical MOSFETs versus trr.

Of course, as suggested before, there would be more parameter to be taken into consideration to define the ruggedness of the MOSFET in ZVS converters, but the only one easily available in data sheet and quite easily measurable is trr.

• There are equipment working for a long time without any problem and, on the other hand, some identical equipment under the same operative conditions fail in a short time: *Only those ones that employ at least a "weak" device could fail.*

• Equipment that, although working at critical light load, have survived for enough time, have a good chance to work forever: *After the failure of the percentage of weaker MOSFETs, the remaining are more rugged.*

• As a rule, an equipment that had a MOSFET failure, after the only replacement of the broken device, works without any other relevant failure: A*s weak devices are only a small percentage of the total population, it is very unlikely, from the statistic point of view, to repair the equipment with another weak device.*

5. Body Diode Experimental Results

5.1. Evaluation of stored charge remaining in the junction

To evaluate this theory a test circuit (see the simplified schematic Fig. 10) was constructed. Its operation is as follows. First we force forward current into the body diode of the Device Under Test (DUT); then we stop the current flow without applying reverse voltage to the junction. Finally, after a measured period of time, reverse voltage is applied across the body diode junction and the remaining reverse recovery charge is measured.

The channel of the DUT can be turned on or off at anytime before or during the test cycle, in order to understand also the effect of gate polarization.

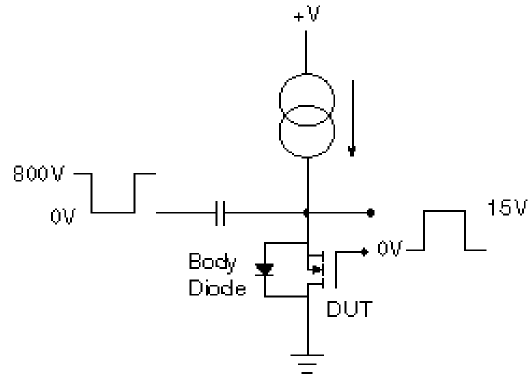


Fig. 10. Simplified schematic.

To determine the capacitance portion of the reverse recovery charge, a preliminary test was conducted without forcing forward current into the body diode and applying a reverse voltage of 800 V at a dv/dt of 1.33 V/nsec. This is defined as the baseline charge, or the amount of charge that represents zero stored charge resulting from forward current flow.

A series of tests were conducted to determine the amount of time required, after the forward current was stopped, for all of the stored charge in the junction to be recombined. The channel remained off during these tests. These tests were conducted forcing 5 A of forward current for 10 μsec and applying a reverse voltage of 800 V at a dv/dt of 1.33 V/nsec. This reverse voltage was first applied at 100nsec after the forward current was turned off. The test was repeated, applying the reverse voltage at a later time, until the baseline reverse recovery charge was reached. These tests were conducted on an APT10026JN, a standard MOSFET and on an APT10025JVFR, a FREDFET (a MOSFET with a fast-recovery body diode). The results are shown in Fig. 11.

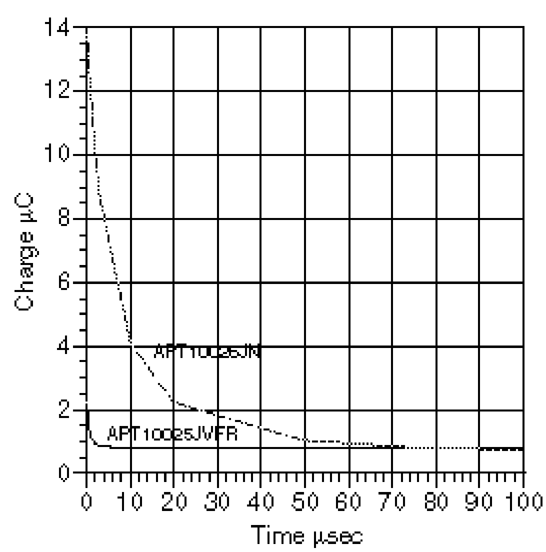


Fig. 11. Stored charge remaining in the junction after the forward current became equal to zero (Vgate = 0 V).

It is clear from Fig. 11 that charge remains in the junction for a time longer than the specified reverse recovery time would indicate. The standard MOSFET was near full recovery at 100 µsec but it took until 200 µsec to fully recover the device.

The FREDFET was near full recovery at 3 µsec and was fully recovered in less than 10 µsec. The FREDFET recovered much faster then the standard MOSFET due to shorter minority carrier lifetime resulting in faster minority carrier recombination.

5.2. Evaluation of the effect of gate bias on stored charge

Another series of tests were conducted to determine the effect of turning on the channel when charge was stored in the junction.

This was obtained by polarizing the gate of the DUT during the forward conduction period. The stored charge was measured at 1 µsec after the forward current was stopped.

As before these tests were conducted forcing 5 A of forward current for 10 µsec and applying a reverse voltage of 800 V at a dv/dt of 1.33 V/nsec. This reverse voltage was applied at 1 µsec after the forward current was turned off.

The first measurement was taken with the channel off. The second measurement (Fig. 12) was taken with the channel being turned on at 1 µsec before the forward current was turned off and turned off at 300 nsec before the reverse voltage was applied.

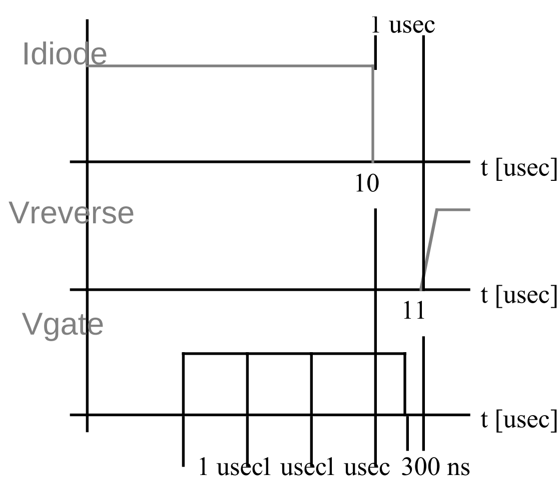


Fig. 12. Timing for measuring the effects of the channel on time on the stored charge remaining in the junction.

Subsequent measurements were taken by increasing the channel on time by 1-µsec steps and turning the channel off always at the same time (300 nsec before the reverse voltage was applied). The results are shown in Fig. 13.

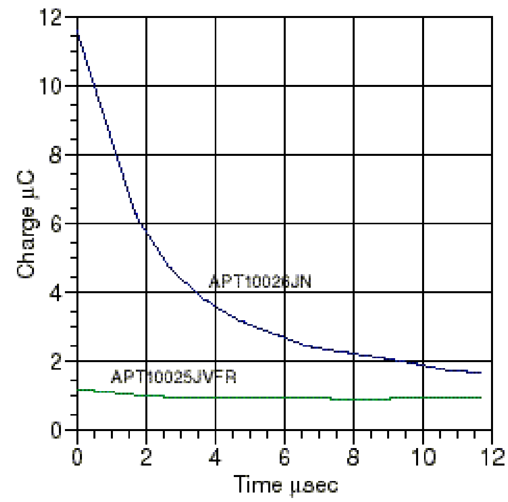


Fig. 13. Stored charge remaining in the junction versus the channel on time (Vgate = 15 V).

It is clear that turning the channel on during the forward conduction of the body diode reduces the amount of stored charge and the sooner the channel is turned on the less the remaining charge is.

However, the standard MOSFET never reaches zero stored charge within a conduction period (10 µsec) even when the channel is turned on before the body diode conducts.

The FREDFET, on the other hand, due to the short minority carrier lifetime, has virtually zero stored charge when the channel is turned on during the last µsec of the of body diode conduction.

5.3. Effect of gate bias on reverse recovery

We prepared a test fixture (Fig. 14) in order to test the reverse recovery behavior of body-diodes. We were not interested so much in an “absolute measurement,” but rather in a “comparative measurement” under the same test conditions.

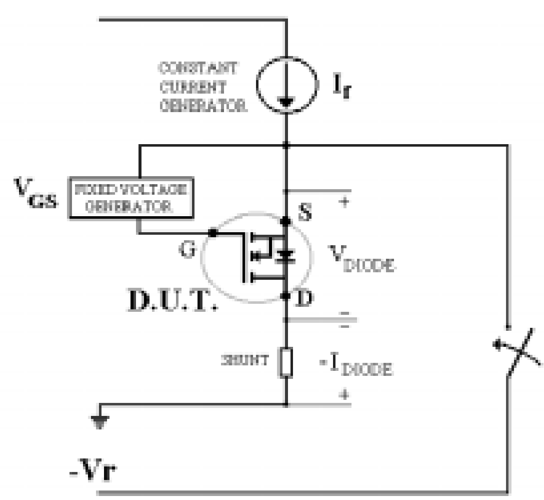


Fig. 14. Test fixture for reverse recovery measurement.

The text fixture is used as follows. A constant DC current I_f is forced across the body diode with a current generator, then we abruptly reverse-polarize the diode junction by means of an external switch connected to the source that applies a negative voltage V_r to the drain. A voltage can be applied to the gate, too. The waveform recorded are the diode current (drain current) and the diode voltage (source-to-drain voltage), see Fig. 15.

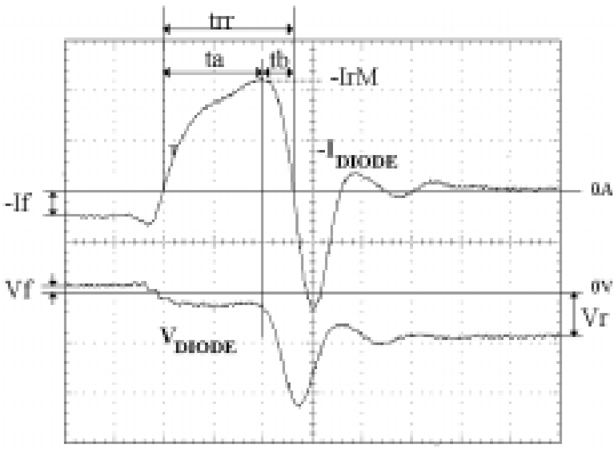


Fig. 15. Example of reverse recovery waveform.

Note that in the figure the diode current is reversed: The negative part of the waveform is the forward current in the diode, the positive part is the recovery current.

In all measurements, the direct current I_f is 2.5 A and the reverse voltage V_r is 1 V. The gate voltage is either 0 V or 3 V.

We tested several MOSFETs. Figs. 16 to 19 show the test results for a standard MOSFET and for a FREDFET.

<ins>STANDARD MOSFET</ins>

As Fig. 16 shows, with Vg = 0 V, tb is large. That means that the charge recombination requires a lot of time. In fact, this interval is completed only when the minority charges that are at some distance from the junction have been swept back to the junction and, in addition, the transition capacitance across the reverse biased junction has been charged to Vr.

When the gate is biased to 3 V (Fig. 17), both trr and Qrr (reverse recovery charge) decrease significantly. Note that 3 V is not high enough to turn on the channel to conduct appreciable current.

<ins>FREDFET:</ins>

In a FREDFET we can observe the same phenomenon, but in this case the reduction in trr is smaller (Figs. 18 and 19).

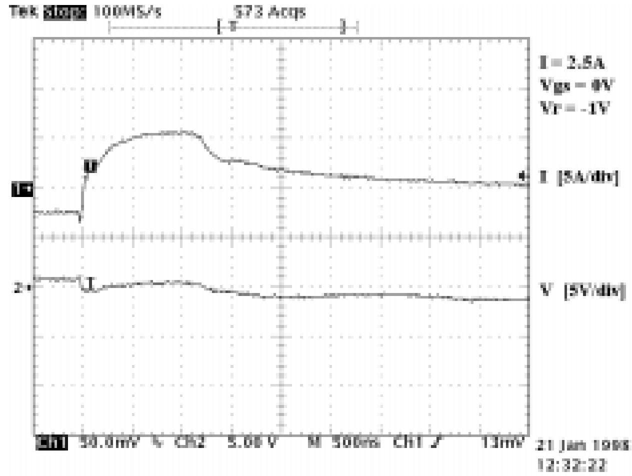


Fig. 16. Reverse recovery of a standard MOSFET. I_f = 2.5 A, V_r = 1 V, Vg = 0 V. Horizontal scale: 500 ns/div.

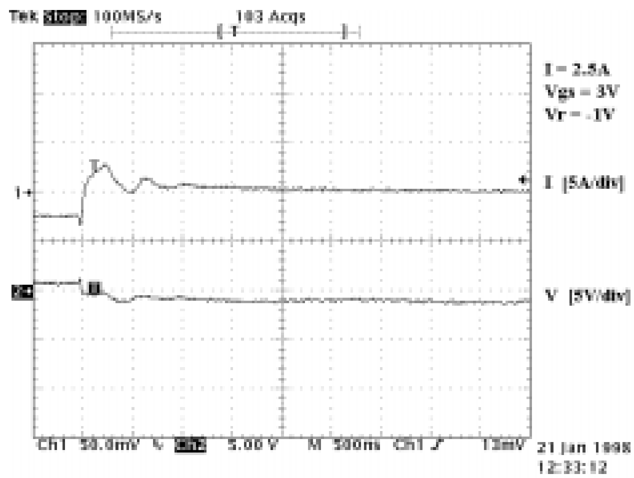


Fig. 17. Reverse recovery of a standard MOSFET. I_f = 2.5 A, V_r = 1 V, Vg = 3 V. Horizontal scale: 500 ns/div.

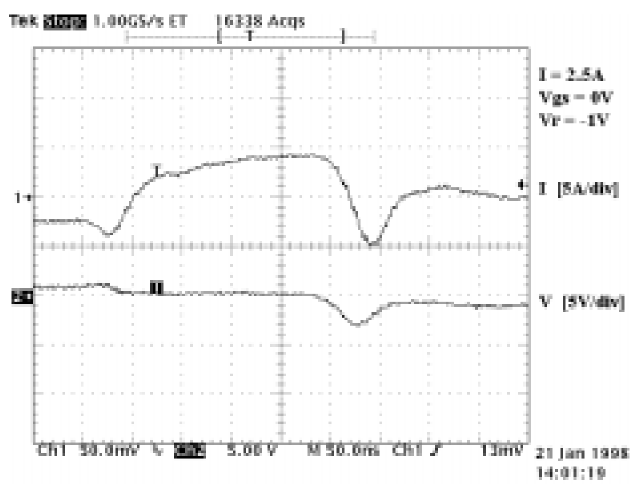


Fig. 18. Reverse recovery of a FREDFET. I_f = 2.5 A, V_r = 1 V, Vg = 0 V. Horizontal scale: 50 ns/div.

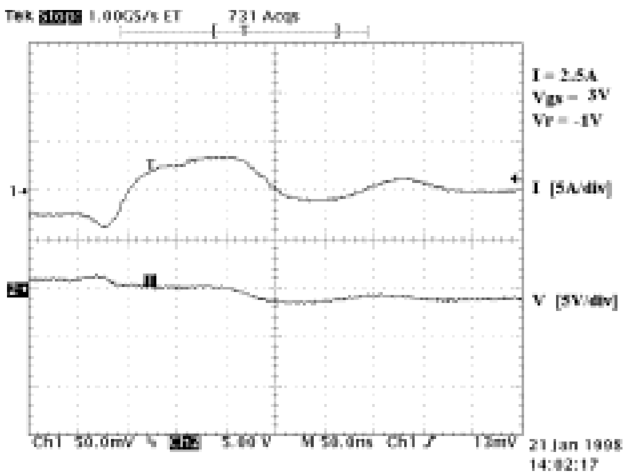


Fig. 19. Reverse recovery of a FREDFET. $I_f = 2.5$ A, $V_r = 1$ V, Vg = 3 V. Horizontal scale: 50 ns/div.

The conclusion of these tests, performed on several MOSFETs from different suppliers, is that a small positive gate bias (close to, but below, the threshold voltage) has a large impact on reverse recovery of the body diode, even though the channel is practically off.

A qualitative explanation of this behavior is that the additional electrons through the channel tend to recombine with the excess holes stored in the N-epi layer, thereby expediting the body-diode recovery.

5.4. Effects of the channel current and channel on time on the stored charge remaining in the junction

Another series of tests was conducted to determine the effect of the channel current on the stored charge by forcing a forward diode current in the DUT, then reversing the polarity of the current (i.e. changing the current from forward diode current to forward channel current of equal magnitude), and maintaining the forward channel current for 2 µsec. The DUT channel was turned on at or before the current reversal and was turned off 2 µsec afterward. Then, 200 ns after the channel was turned off, we applied a reverse diode voltage of 800 V at a dv/dt of 1.33 V/nsec and measured the stored charge. These tests were conducted with 2 A and 9 A of forward diode current flowing for 5µsec. The first measurement was taken with the channel being off. The second measurement was taken with the channel being turned on 1 µsec before the current reversal. Subsequent measurements were taken by turning the channel on earlier in 1-µsec increments, keeping the channel on for 2 µsec after the current reversal and turning the channel off at 200 nsec before the reverse voltage was applied. The results are shown in Fig. 20. It is clear from the figure that although the 9 A forward diode current resulted in a higher initial stored charge, than the 2 A forward diode current, it was removed more efficiently when the channel was turned on. The reason for this is the higher voltage drop across the channel, resulting from the higher current, will sweep more charge across the

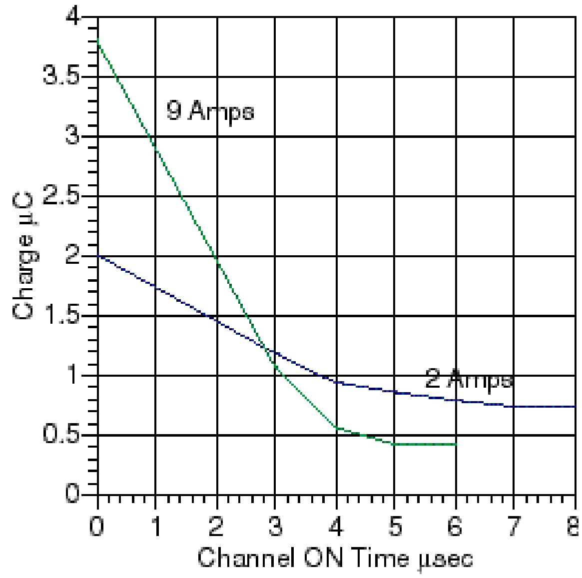


Fig. 20. Stored charge remaining in the junction versus channel on time with 2 A and 9 A forward current flowing.

junction. This explains why failures occurred during lower power operation. There is more charge remaining in the diode when high voltage is applied making the diode more susceptible to second breakdown.

6. Technical Solutions to Improve the Equipment Reliability

In order to improve the ruggedness of the device and consequently the reliability of the equipment several solutions were investigated.

6.1. Using FREDFETs

MOSFETs with faster reverse recovery time of the body diode show more ruggedness when working in a ZVS converter. Nowadays MOSFETs are available with body-diode reverse recovery times which are about one-tenth of the recovery times of devices from only a few years ago. This is especially true for high voltage (800-V and 1000-V) devices. Our experience shows that the long term failure rate of ZVS converters built with 800-V or 1000-V FREDFETs is much smaller than that of ZVS converters built with standard high voltage MOSFETs.

We also considered three other possibilities, all of which require some modification of the ZVS converter circuitry:

6.2. Converter modifications

The goal of the modifications is to avoid using the body diode, or at least to minimize the consequences. Three possibilities were considered:

1) The conduction of the body diode can be prevented by adding current-steering diodes (a Schottky diode in series with the MOSFET drain, along with an ultrafast antiparallel by-pass diode), see Fig. 21.

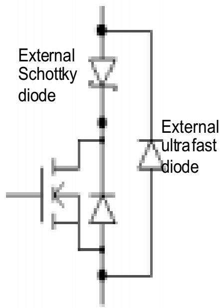


Fig. 21. Avoiding body-diode conduction with current-steering diodes.

In this case the body diode never conducts. However, this solution is quite expensive and complex; also the series diode increases the conduction losses.

2) We observed that at output loads above 25% of full load (that means high current in the ZVS converter and consequently high voltage drop across $RDS_{(on)}$ during the on time) the failure rate is negligible. It is expected that the failure rate will be reduced at light load if we keep the drain-to-source voltage high enough to force the complete recombination of the minority charges before the end of the on time. That can be achieved by adding an antisaturation circuit. Fig. 22 shows the simplified schematic of such a circuit.

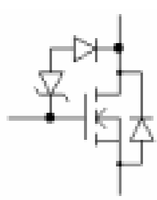


Fig. 22. Simplified antisaturation circuit.

Using a zener diode with a proper voltage, at light load the MOSFET is forced to operate in its linear region. In that region the drain-to-source voltage is not related anymore to the current flowing in the MOSFET, rather it is fixed and is equal to the difference between the drive voltage and the sum of the zener plus rectifier diode voltages. The zener voltage should be chosen to force the drain-to-source voltage to be at least as high as it was at 25% of full load.

At high load, the voltage drop across $RDS_{(on)}$ is higher than the fixed value set by the drive, zener and rectifier voltages, and the behavior is as usual.

The main problems of this solution are:

- zener voltage tolerance and temperature dependence,
- power consumption of the MOSFET driver,
- circuit complexity and cost.

3) Similar to the antisaturation circuit, the third solution only prevents the conduction of the body diode when that might cause problem, i.e. at light load. This turns out to be an effective and simple solution, without further increasing the complexity of the converter.

A closer look at the basic operation of the soft-switching full-bridge converter (stages 2 to 4) reveals that the body diode can conduct only if the drain-to-source voltage reaches zero before the switch is turned on (Fig. 23).

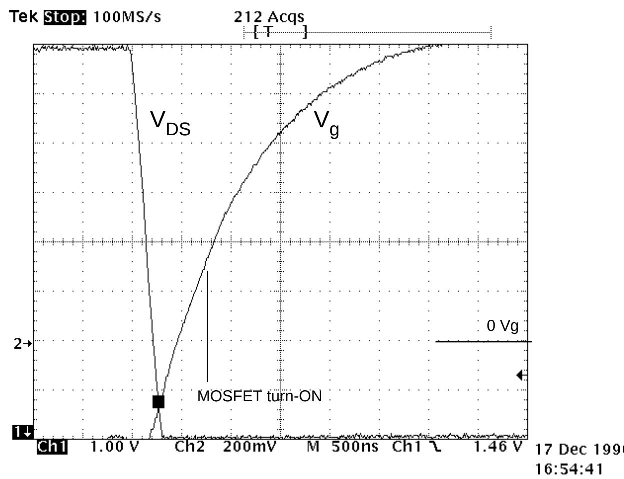


Fig. 23. Standard behavior at light load. Io = 10 A, V_{DS}: 100 V/div., Vg: 2 V/div., 500 ns/div.

If we force the MOSFET to turn on before its drain-to-source voltage reaches zero (Fig. 24), the channel of the MOSFET will be placed in parallel with the body diode before current starts flowing in the device. That prevents the flow of current in the body diode. In fact if the voltage drop across the $RDS_{(on)}$ is less than the threshold voltage of the body diode, the current will flow only in the channel. This is exactly what happens at light load.

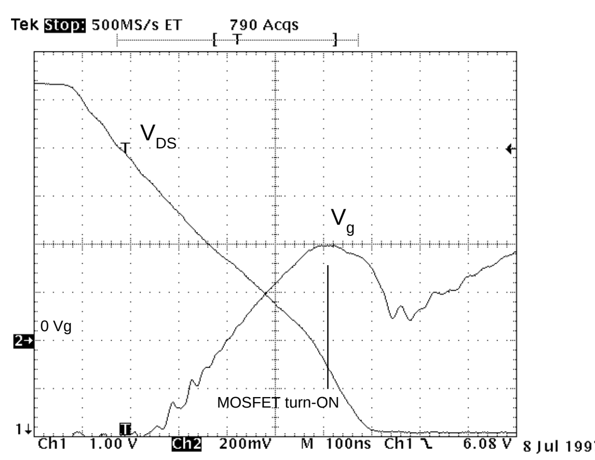


Fig. 24. Improved behavior at light load. Io = 10 A, V_{DS}:

100 V/div., Vg: 2 V/div., 100 ns/div.

We can obtain the turn-on of the MOSFET before its drain-to-source voltage reaches zero either by (1) shortening the turn-on delay or (2) reducing the dv/dt of the drain-to-source voltage. Both can be easily performed: The first one requires changing the value of a resistance in the control board that sets the proper delay between the switches, the second one requires increasing the value of the external snubber capacitor placed in parallel with each switch. (Those capacitors are normally required for reducing the turn-off loss; they slow down the rise of the drain-to-source voltage, so that the current turn-off completes with negligible voltage across the MOSFET.)

Note that by turning on the MOSFET with nonzero voltage we lose perfect zero-voltage switching. This however, is not a problem since it happens only at light load, where a slight drop in efficiency can be easily tolerated. Anyway, we have a "quasi zero-voltage-switching" behavior, as the drain-to-source voltage has enough time to approach zero before the MOSFET turns on. Since the turn-on loss is proportional to the square of the voltage across the snubber capacitor, the quasi zero-voltage-switching is sufficient to keep the turn-on loss at an acceptably low level.

If the load current increases, the slope of the drain-to-source voltage also increases and since the turn-on delay is fixed, the drain-to-source voltage gets closer to zero at the MOSFET turn-on instant (Fig. 25).

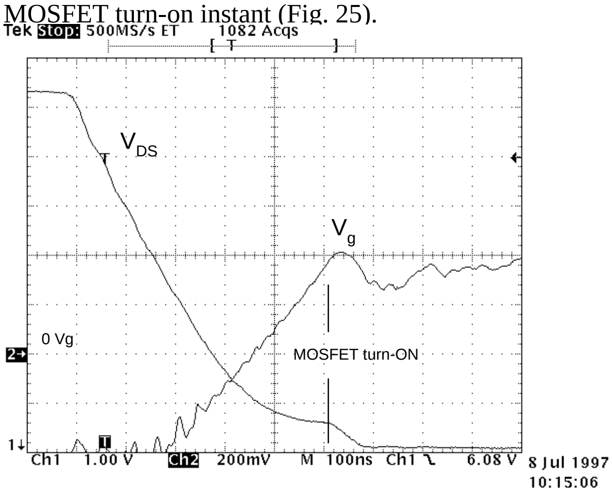
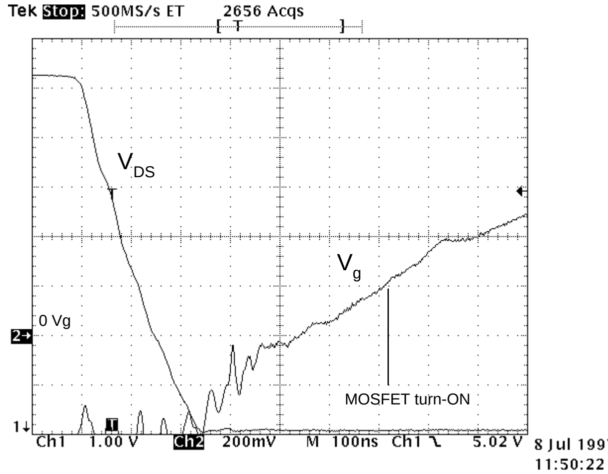


Fig. 25. Quasi zero-voltage switching at medium load. Io = 30 A, V_{DS}: 100 V/div., Vg: 2 V/div., 100 ns/div.

Above a certain load, the drain-to-source voltage will reach zero before the MOSFET turns on. In this case the converter operates with perfect zero-voltage switching (Fig. 26).

The application of this solution has dramatically lowered the MOSFET failure rate, without adding any further complexity to the circuitry and layout.

In mass production this solution was used in combination with FREDFET devices. The effectiveness of the combined solution was proved by a statistical analysis performed on



more than 500 equipment that have been

Fig. 26. Zero-voltage-switching at high load. Io = 75 A, V_{DS}: 100 V/div., Vg: 2 V/div., 100 ns/div.

working for more than 3 million hours in total. That analysis confirmed that the failure rate of the MOSFETs in the ZVS converter is now negligible.

Summary

This paper attempts to explain some until now unexplainable failures which were observed in zero-voltage switching converters (in particular in the PWM phase-shift full-bridge converter) employing high-voltage MOSFETs. The main failure mechanism of MOSFETs is the second breakdown of the parasitic bipolar transistor. A second breakdown can happen when the parasitic bipolar transistor is activated by the residual charge left over from the conduction of the body diode. A common belief is that in zero-voltage-switching converters, where the conducting body diode is normally shunted out with the resistance of the channel, no charge remains in the device after the conducting period. As explained in the paper and also as experimental data show, this is not true: The charge remaining in standard MOSFET devices can be substantial and often is sufficient to trigger the second breakdown. On the other hand, the charge remaining in MOSFETs with fast-recovery body diodes (FREDFETs) is small enough that the second breakdown problem almost vanishes. Complete protection against second breakdown can be achieved by combining FREDFETs with a small modification of the converter. That modification is to increase the snubber capacitances placed across the MOSFETs such that the channel turns on before the drain-source voltage would become zero and the body diode would start conducting. As discussed in the paper, this can be done without any detrimental effect on the full-load converter efficiency, since, as shown in the paper, the early turn-on of the channel is necessary only at light load. While a small efficiency reduction can be expected at light load due to the loss of perfect zero-voltage switching, the full-load efficiency is expected to increase somewhat due to the reduction of the turn-off losses brought about by the larger snubber capacitances. A statistical analysis of the failure of power supplies installed in the field proves that the combination of FREDFETs and the increased snubber capacitances practically eliminates the MOSFET failures triggered by the stored charge of the body diode.

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References

[1] Richard Redl, Laszlo Balogh, and Nathan O. Sokal, "A Novel Soft Switching Full Bridge DC/DC Converter: Analysis, Design Considerations and Experimental Results at 1.5 kW, 100 kHz, " *PESC '90 Record*, pp. 162-172.

[2] Bill Andreycak, "Designing a Phase Shifted Zero Voltage Transition (ZVT) Converter," Topic 3 in the Unitrode Power Supply Design Seminar Manual, SEM-900, 1993, Unitrode Corporation.

[3] Bill Andreycak, "Design Review: 500W, 40W/in^3 Phase Shifted ZVT Power Converter," Topic 4 in the Unitrode Power Supply Design Seminar Manual, SEM-900, 1993, Unitrode Corporation.

[4] Bill Andreycak, "Phase Shifted, Zero Voltage Transition Design Considerations and the UC3875 PWM Controller," Unitrode Application Note U-136A.

[5] Brian Pelly, "The Do's and Dont's of Using the Power HEXFET," International Rectifier Application Note 936A.

[6] "Avalanche and dv/dt Limitation of the Power MOSFET," Chapter 5 in the Motorola TMOS Power MOSFET Transistor Device Data Book, DL 135/D.

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