AN-557 Optimizing the Ultra-Fast POWERplanar Rectifier Diode for Switching Power Supplies
INTRODUCTION
A key device in all high voltage AC-DC power supplies is the ultrafast, reverse recovery rectifier diode. These diodes (D1 and D2 in Figure 1) not only play a major role in power supply efficiency but also can be major contributors to circuit electromagnetic interference (EMI) and even cause transistor failure if they are not selected correctly. One would assume that by now, this rectifier diode should approximate the behavior of an ideal switch, i.e., zero on-state voltage, no reverse leakage current and instantaneous turn-on. At first glance, the design of this single pn-junction device would appear to be quite straightforward but a review of the device equations reveals that many compromises must be made to optimize its performance. An understanding of these tradeoffs will allow the circuit designer to select the most appropriate rectifier diode.

FIGURE 1a. Buck Regulator to Step-Down Input Voltage $V_{IN}$

Consider how the non-ideal behavior of rectifier D2 affects the circuit performance of the buck regulator in Figure 1a. The solid lines in Figure 2a depict the switching behavior of the transistor switch and rectifier in comparison to the waveforms (dashed lines) that represent an ideal rectifier. There are four differences between the two cases:

1. The most significant difference is that the peak collector current of the transistor switch ($I_T$ in Figure 2a) at the end of turn-on (time $t_2$) has been increased by the magnitude of the peak reverse recovery current of the rectifier ($I_{REC}$). Correspondingly, the peak power dissipation within the transistor has increased from $P_T$ to $P_T'$ as shown in Figure 2c.

2. The maximum transistor voltage $V_T$ at turn-off ($t_4$ to $t_6$ in Figure 2a) has been increased by the dynamic voltage drop of the rectifier during turn-on. Since buck regulators generally run at low voltages, this increase has a minimal effect. However, it is more significant in the forward converter circuit of Figure 1b and in bridge circuits operating from high bus voltages where the voltage margins cannot be as generous.

3. Since the rectifier is not ideal, its power dissipation consists of the following components:
   a. Conduction loss ($V_F \times I_F$) during the on-time.
   b. Turn-off loss during time $t_2 \sim t_3$ and turn-on loss during time $t_5 \sim t_6$ (Figure 2d).
   c. Reverse blocking loss ($V_R \times I_R$) during period $t_3 \sim t_5$.

4. The rectifier regains its reverse blocking capability at time $t_2$. A "snappy" rectifier that quickly turns off $I_{REC}$ will contribute much more EMI than a "soft", fast recovery rectifier.

A better transistor switch will intensify rather than improve the shortcomings of the fast recovery rectifier, so it is necessary to consider more fully the conduction and switching behavior of the rectifier diode.
FIGURE 2. Transistor and Rectifier Voltage and Current Waveforms for the Buck Regulator in Figure 1a

a) Transistor and Rectifier Voltage Waveforms
b) Transistor and Rectifier Current Waveforms
c) Transistor Power Dissipation
d) Rectifier Power Dissipation

POWER LOSSES IN THE ULTRA-FAST RECTIFIER DIODE

Consider the idealized rectifier current and voltage waveforms in Figure 3 for a 50 kHz buck regulator. Power dissipation within the rectifier for a 50% duty factor is:

\[ P = P_{\text{conduction}} + P_{\text{blocking}} + P_{\text{reverse recovery}} \]

\[ P = \frac{1}{2} \left[ (V_F + V_R + V_{RM(REC)}) t_f \right] \]

Typical values for a 200V, 8A rectifier are:

\[ f = 50 \text{ kHz} \]
\[ I_R = 1 \text{ mA} \]
\[ V_F = 0.9 \text{ V} \]
\[ t_b = 25 \text{ ns (assuming } t_b = t_r/2) \]
\[ I_F = 8 \text{ A} \]
\[ V_R = 50 \text{ V} \]
\[ I_{R(REC)} = 2.0 \text{ A} \]
\[ V_{RM} = 200 \text{ V} \]

\[ P = \frac{1}{2} \left[ (8 \text{ A})(0.9 \text{ V}) + (50 \text{ V})(1 \text{ mA}) \right] \]
\[ + (200 \text{ V})(2 \text{ A})(25 \text{ ns}) (50 \text{ kHz}) \]
\[ P = 3.6W + 0.025W + 0.5W = 4.125W \]
Planar passivation techniques have reduced surface leakage currents ($I_R$) to a negligible amount so that the principle reverse leakage current is recombination current in the space charge region. Some of the many methods to control minority carrier lifetimes are electron or neutron irradiation and gold or platinum diffusion, each with its own advantages and disadvantages. For 200V, ultrafast recovery rectifiers, gold diffusion still represents the best compromise between speed, $V_F, I_R$ and "soft" recovery.

A drawback to gold diffusion is its relatively high reverse leakage current. It should be pointed out that the reliability of the gold-diffused product is the same as other rectifiers (all other factors being equal), since this leakage current is a bulk and not a surface phenomenon. Experimental leakage test results have been plotted in Figure 8 for the National Semiconductor 8A and 16A series of rectifiers (FRP820 and FRP1620 respectively) at 100°C, 125°C and 150°C. These points indicate that the low current injection level lifetime ranges from 20 ns–30 ns and is relatively independent of $T_J$. Since reliability design guidelines specify that the rectifiers be operated at one-half their voltage rating and 25°C–50°C below their maximum junction temperature, the expected leakage currents in well designed power supplies will run less than 1 mA.

**REVERSE RECOVERY LOSSES**

All pn-junction rectifiers, operating in the forward direction, store charge in the form of excess minority carriers. The amount of stored charge is proportional to the magnitude of the forward current. The process by which a rectifier diode is brought out of conduction and returned to its block state is called commutation. Figure 7 shows an expanded view of current commutation, also called reverse recovery. Starting at time $t_d$, the rectifier is switched from its forward conducting state at a specified current ramp rate ($-dI/dt$). The current ramp rate will be determined by the external circuit ($E/L$) or the turn-on time of a transistor switch. During the time $t_1$–$t_2$, the store charge within the rectifier is able to supply more current than the circuit requires, so that the rectifier behaves like a short circuit. Stored charge is depleted both by the reverse recovery current and recombination within the rectifier. Eventually the stored charge dwindles to the point that a depletion region around the junction starts to grow, allowing the rectifier to regain its reverse blocking voltage capability ($t_2$). From a circuit-design standpoint, the most important parameters are the peak reverse recovery current and "S", the softness factor. A "snappy" rectifier will produce a large amplitude voltage transient and contribute significantly to electro-magnetic interference. Figure 8 illustrates the actual reverse recovery of two rectifier diodes. The peak voltage of the snappy rectifier is 175V compared to 142V peak for the FRP820, the higher voltage resulting from both the higher $I_{(REC)}$ and the fact that the reverse recovery current decays to zero in a shorter time.
FIGURE 7. Expanded View of Current Commutation in a Rectifier Diode

The relative snappiness of a rectifier may be defined quantitatively by dividing the reverse recovery time \( t_{rr} \) into two subperiods, \( t_a \) and \( t_b \), as shown in Figure 7. The softness factor “S” is simply the ratio \( t_b/t_a \). A rectifier with a low value S factor will be more likely to produce dangerous voltage transients, but it will also dissipate less reverse recovery energy than a high S factor rectifier. A reasonable compromise between these two conflicting constraints would be to design a rectifier with \( S \approx 1 \) (\( t_a = t_b \)). The S factors of the FRP820 rectifier and the competitive device in Figure 8 are 0.55 and 0.31 respectively.

Only recently has it become possible to model the ramp recovery in p-i-n rectifiers (References 2, 3) and the following equations have proved useful in predicting reverse recovery parameters.

\[
\begin{align*}
t_{rr} &= \frac{W_i}{\tau Da} \\
S &= \frac{W_i}{A Da} \tau \\
I_{R(REC)} &= \left( \frac{dI_F}{dt} \right) \tau \left( 1 + \frac{W_i}{8 Da} \right) - 1 \\
Q_{R(REC)} &= 0.5 \tau^2 \left( \frac{dI_F}{dt} \right)
\end{align*}
\]

where:

- \( \tau \) = minority carrier lifetime
- \( W_i \) = epitaxial thickness
- \( Da \) = ambipolar diffusion constant

The blocking voltage rating of the rectifier primarily determines \( W_i \); but for a given \( W_i \), note that a short minority lifetime not only decreases \( I_{R(REC)} \) but happily increases S. These two key parameters are plotted as a function of minority carrier lifetime in Figure 9 for \( dI_F/dt = 100 \text{ A/} \mu\text{s} \) and \( T_J = 25^\circ\text{C} \). As has been noted before, the minority carrier lifetime had been targeted for 20 ns–50 ns to minimize \( V_F \) and this choice has resulted in a typical value of \( S \approx 0.65 \) and \( I_{R(REC)} \approx 1.5A \).

FIGURE 8. Comparison of Reverse Recovery of the FRP820 Series Rectifier to a Snappy Rectifier

REVERSE RECOVERY CHARACTERIZATION

Figures 10–13 plot \( Q_{R(REC)} \), \( I_{R(REC)} \), \( t_{rr} \), and \( S \) versus \( dI_F/dt \) for the FRP1600 series of rectifiers and typical use conditions of \( I_F = 16A \) and \( V_R = 200V \) and for two different junction temperatures of 25°C and 125°C. Theory not only predicts, but it has also been experimentally verified, that these parameters are relatively independent of \( I_F \) so only one value of the latter suffices. Any three of the four Figures 10–13 completely specifies the reverse recovery behavior of the rectifier. Since \( S \) and \( T_{rr} \) vary the least over the plotting \( dI_F/dt \) range, it is convenient to formulate reverse recovery energy loss \( P \) in microwatts in terms of the circuit parameters \( V_R \) and \( dI_F/dt \):

\[
P = \frac{V_R}{2S} \left( \frac{dI_F}{dt} \right)^2 (Sf)^2 \cdot 10^{-3} (\mu\text{W})
\]

where:

- \( V_R \) = peak reverse voltage
- \( dI_F/dt \) = ramp rate (A/\( \mu\text{s} \))
- \( f \) = operating frequency (kHz)
Example: Calculate the reverse recovery power loss for the FRP1620 rectifier running at:
\( I_F = 16A \), \( V_R = 100V \),
\( \frac{dI_F}{dt} = 100 A/\mu s \), \( T_J = 125^\circ C \), \( f = 75 \text{ kHz} \).

From Figures 12 and 13, \( t_{rr} = 56\text{ ns} \) and \( S = 0.29 \). Substituting these values in the above equation:
\[
P = \left( 100V \right) \left( 100 A/\mu s \right) \left( 75 \text{ kHz} \right) \left[ \frac{0.29 \times (56 \text{ ns})}{1 + 0.29} \right]^{2} \times 10^{-3} \mu W
\]
\[
P = 0.205 W
\]

There are may ways to shape the reverse recovery voltage spike. The most simple and still most popular is the RC snubber circuit connected across the primary of the transformer in Figure 1b. This serves the dual purpose of suppressing voltage ringing and EMI due to the switching action of both the transistor and rectifier. William McMurray has shown how to design an RC snubber to minimize voltage transients and/or \( \frac{dv}{dt} \) ramps just due to the diode reverse recovery current (Reference 4) and also how to de-
sign snubbers to minimize transistor power dissipation (Reference 5). But to date, because the RC snubber plays a major role in reducing EMI, its design tends to be empirical rather than theoretical.

CONCLUSION
This application note has pointed out the major considerations in designing an ultrafast reverse recovery rectifier and shown that the control of minority carrier lifetime is the key in arriving at an optimum device. Because the diode contributes to EMI, its reverse recovery behavior must be carefully controlled and characterized in order to guarantee similar performance from lot to lot.

REFERENCES
3. C. M. Hu, Private Communication

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