SGS-THOMSON MICROELECTRONICS

APPLICATION NOTE

HIGH VOLTAGE TRANSISTORS WITH POWER MOS EMITTER SWITCHING

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INTRODUCTION

This paper summarizes the results of an investigation carried out on power devices with both MOS and BIPOLAR parts working together in the same circuit. The "emitter drive" configuration was considered, with switching power supply applications in mind.

The devices used are :

Power MOS	: SGSP321, IRFZ20
Bipolar transistors	: BUV48, BU508A
Ultrafast bipolar transistors	: SGS463
(Hollow Emitter)	: SGS443
Fast darlingtons	: SGSD00055, BU810

In the case of flyback switching power supplies a practical example is also described.

CIRCUIT DESCRIPTION

The term "emitter switching" describes a circuit configuration where a low voltage transistor (MOS or Bipolar) switches off the emitter current of a high voltage transistor, and consequently the transistor itself.

This configuration combines the fast switching of a low voltage device with the high power switching of a high voltage device, since :

high current x high voltage = high power switching.

The combination of a high voltage bipolar and a low voltage Power MOS is preferable due to the high switching speed and the low driving energy of the combined power switch.

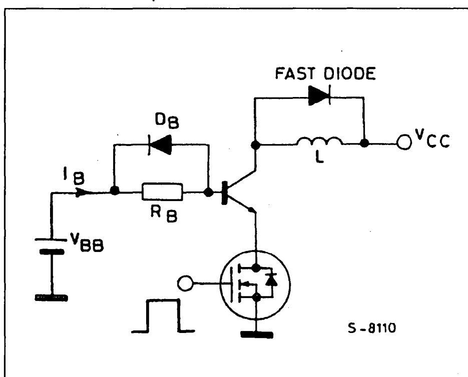
The base of the high voltage bipolar device is driven by a constant voltage source. The energy dissipated to drive the high voltage bipolar device depends on the losses that the forward bias current I_{B1} generates in the resistance in series with R_B, $I_{B1}^2 \cdot R_B \cdot t$. This power dissipation can only be reduced by using high gain transistors or Darlingtons.

(see fig. 1).

The diode in series with the base serves to clamp the base over voltage at turn-off.

The two transistor stage is driven by the gate of the low voltage Power MOS. Very low driving energies, about 180nJ per cycle, are involved in the charging of the input capacitances.

Figure 1 : The Basic Circuit used for the Evaluation of the Emitter Switching System. The Base Drive Circuit used is shown for Comparison.



Consequently the stage can be directly driven by the output of suitable linear integrated circuits.

The possibily of direct driving by an IC output together with the excellent switching speed make this configuration extremely suitable for switching power supplies at frequencies of 50kHz or higher.

CIRCUIT OPERATION

As we have seen, the forward base current I_{B1} is fixed by the external circuitry :

$$I_{B1} = \frac{V_{BB} - V_{BEsat} - V_{DSon}}{R_B}$$

The collector current instead depends on the load, and in general, varies with the time.

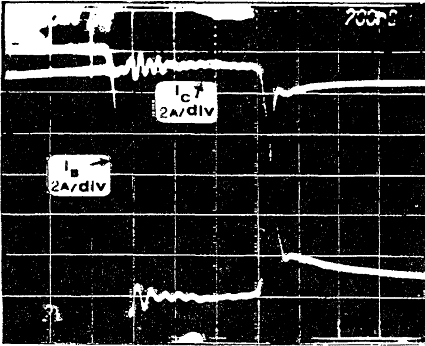
The turn-on and turn-off phases can be analysed separately.

TURN-OFF

When the driving signal to the Power MOS is low the drain current is interrupted and the emitter current of the high voltage bipolar falls to zero. The emitter reaches the base voltage and will not carry any more current. As a result the collector current can only flow through the base, becoming a reverse base current that depletes the base to collector junc-

tion. This reverse base current I_{B2}, from the moment when the emitter current disappears, coincides with the collector current. See photo 1. The stored charge is removed in a typically very violent, and consequently rapid manner.

Photo 1 : Base and Collector Current at Turn-off.



As a result the storage time is substantially reduced.

The fall time, which is related to the recombination under the emitter, is also generally reduced.

Typical values for the fall and storage time of the SGS-Thomson devices used in the test are shown in table 1, for both emitter and the base drive circuits.

Table 1 : Typical tf and ts on Inductive Load.

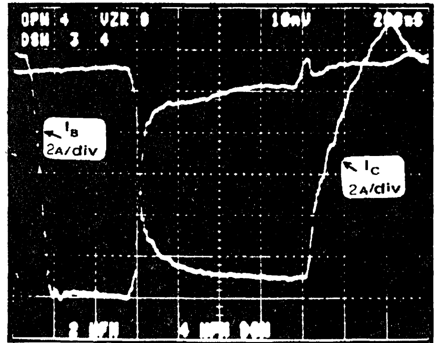
Device	$I_{C(A)}$	Emitter switching		Base Switching	
		tstorage	tfall	tstorage	tfall
BUX48	10	500ns	100ns	2μs	200ns
BU508A	5	800ns	300ns	6μs	400ns
SGSD00055	10	400ns	100ns	1.2μs	100ns
BU810	5	300ns	150ns	800ns	150ns
SGSD00035	10	300ns	50ns	800ns	50ns
SGSD00039	5	300ns	40ns	700ns	50ns

TURN-ON

When the Power MOS is the on state, the bipolar device also starts conducting. The dynamic behaviour (see photo 2) does not differ in any substantial way from the usual case of the base drive.

The dynamic saturation transient $V_{CEsat\ dyn}$ is also practically the same with a base drive as with an emitter drive. The collector current, when the collector load is the primary winding of a switching transformer, can vary according to two possibilities, (see fig. 2).

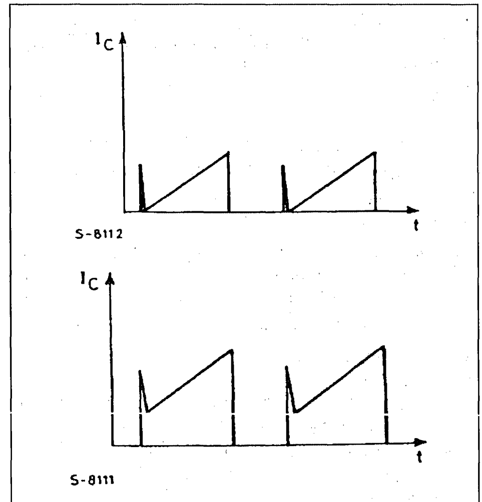
Photo 2 : Base and Collector Current at Turn-on.



a) After the initial peak due to the recovery of the diode present on the secondary winding, the collector current increases linearly starting from zero

b) After the same initial peak, the collector current increases linearly starting from the value memorized in the magnetic circuit at the end of the previous cycle.

Figure 2 : Collector Current Waveforms with Varying Load.



REVERSE BIAS SAFE OPERATING AREA

A problem that occurs in bipolar transistors is damage caused by "current crowding".

Fig. 3a illustrates current flowing in a typical bipolar device. Fig. 3b shows how, when the device is

turned off and the current begings to die away, the current focuses with a high concentration under the emitter. This high current density can damage or destroy the transistor.

Figure 3a.

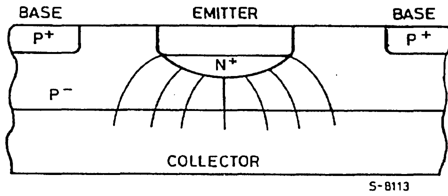
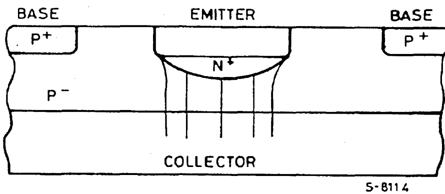


Figure 3b.



The energy dissipated within a bipolar power transistor at turn-off can be found graphically from a plot of I_C versus V_{CE} at turn-off. Three cases are shown in fig. 4a, b and c. The shaded area is proportional to the energy that is dissipated in the device during turn-off.

Consequently turn-off times affect the SOA of the device, (fig. 5b). These problems can be overcome using emitter switching.

Figure 4a : Slow Turn-off. No Crowding but High Average Heating.

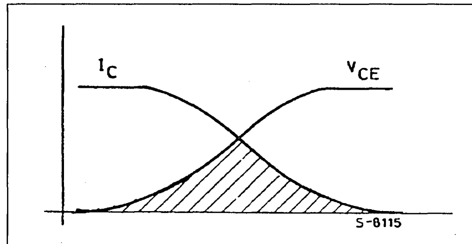


Figure 4b : Fast Turn-off. Crowding with Low Average Heating but Possible High Peak Power.

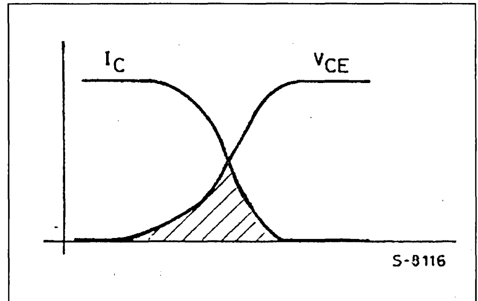
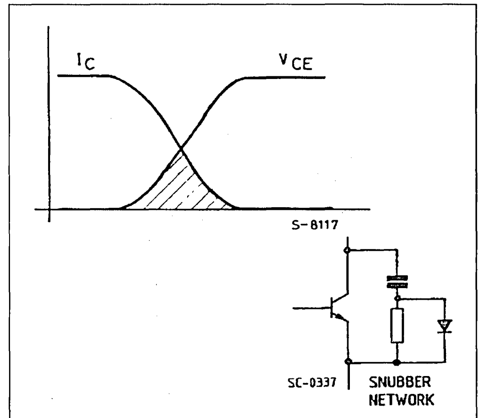


Figure 4c : Fast Turn-off (with V_{CE} delayed by snubber network).



The way the stored charge is swept away in the high voltage bipolar device when it is driven by the emitter, produces some interesting consequences.

The stored charges are evacuated through the base contact when the emitter current is zeroed and not later than a few tens of nanoseconds after the beginning of the storage interval. Consequently, during the turn-off, no charge is injected from the emitter into the base. Although the reverse base current is quite relevant, no focusing of the current in the centre of the emitter fingers takes place.

The bipolar device therefore exhibits an energy absorbing ability at the turn-off RBSOA that is substantially higher than if a normal base drive were used. With a base drive the emitter would inject charges and the voltage drop across the distributed base resistance would induce the "emitter crowding" phenomenon.

The practical evidence for all the transistors investigated (BUV48, BU508A, SGSF463) shows that the reverse bias operating area (RBSOA) extends right up to the BV_{CES} ! (see fig. 5).

This extreme effect is unfortunately much less pronunced when using fast Darlingtons. The higher complexity of the charge extraction mechanism and the charge injection from the emitter into the base in the driver transistor imply that the RBSOA extension is almost irrelevant.

Figure 5a : Reverse Bias Safe Operating Area.

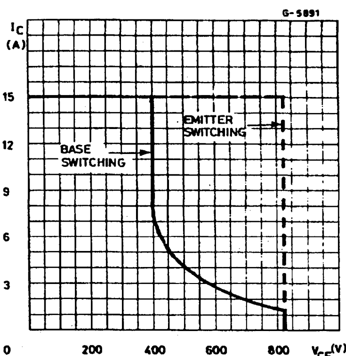
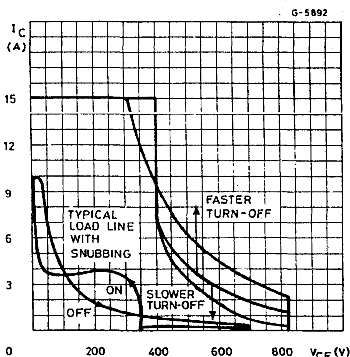


Figure 5b : How Reverse Bias Safe Operating Area Changes for :
i) slow turn-off.
ii) fast turn-off.



A POSSIBLE APPLICATION

A possible application of the "emitter switching" configuration is shown in figure 6, where a switching power supply operating in a "flyback" mode has been implemented.

The basic criteria used in choosing the value of the circuit elements are given below. The purpose of the study was to demonstrate the feasibility and to evaluate the advantages. Exact circuit element values can be further optimized, especially in the case of the transformer.

The power source in the mains singlephase, 220V a.c., and the switching frequency can be set to 50kHz or more.

The devices used were :

Q1 : fast Darlingtons with BV_{CES} 1000V for 110V line
- SGSBU810 for current up to 5A
- SGSD0055 for current above 5A

Fast transistor with $BV_{CES} \geq 800V$ for 220V line
- SGSF443 for currents up to 5A
- SGSF463 for current up to 10A

Q2 : Low voltage POWER MOS ($BV_{DSS} = 50V$)
- SGSP321/IRFZ20 for currents up to 10A

Q3 : High voltage, low current POWER MOS $V_{(BR)\ DSS} \leq$ 450V)

Control IC : UC3842

DZ2 : Zener diode 2W/20V

D1 : 25V diode, with I_c peak rating as high as 10A for 500ns

C6 : Electrolytic capacitor, 100µF, 25V. It absorbs possible variations of V_{BB}.

R3 : Resistor setting the forward bias base current of the Darlington :

$$R3 = \frac{V_{CE} - V_{BEsat} - V_{DSon} - R_7\ I_D}{I_{B1}}$$

Its power rating must exceed $R3 \cdot I_B^2 \cdot t$ (in practice 3W)

R7 : Shunt resistor to sense the switch current. The over current $I_{S\ max}$ protection is set according to

$$R7 = \frac{1V}{I_{Smax}}$$

C4, R6 : RC network, filtering the disturbances induced by the switching transients on the I_{Smax} protection input.

C3, R5 : RC network, setting the switching frequency and the maximum duty cycle, according to the UC3842 data sheet.
$t_{charge} = 0.55\ R_5\ C_3$
$t_{discharge} =$
$R_5 \times C_3 \ln [(6.3\ R_5 - 2.7)/(6.3\ R_5 - 4)]$
$f = 1/(t_c + t_d)$

R8, R9 : Resistive divider of the feedback voltage from a secondary sense winding, rectified by D5 and C5. The divided voltage is compared by the control IC to an internal reference of 2.5V.

C2, R4 : Compensating network in the error amplifier of the feed-back voltage.

R1 : Resistor biasing the Q3 gate (1.2M, 1/4W)

R2 : Resistor that limits the inrush current through the POWER MOS, Q3, at the turn-off (1.2, 2W)

D4 : Fast recovery diode
Its voltage/current ratings depend on the particular secondary winding it rectifies.

D5 : Low current/low voltage diode

D3, R10, C8 : Snubber network (fig. 6 shows just one of the possible configurations).

$$C8 = \frac{L_d\ I_c^2}{V_{DS}^2}$$

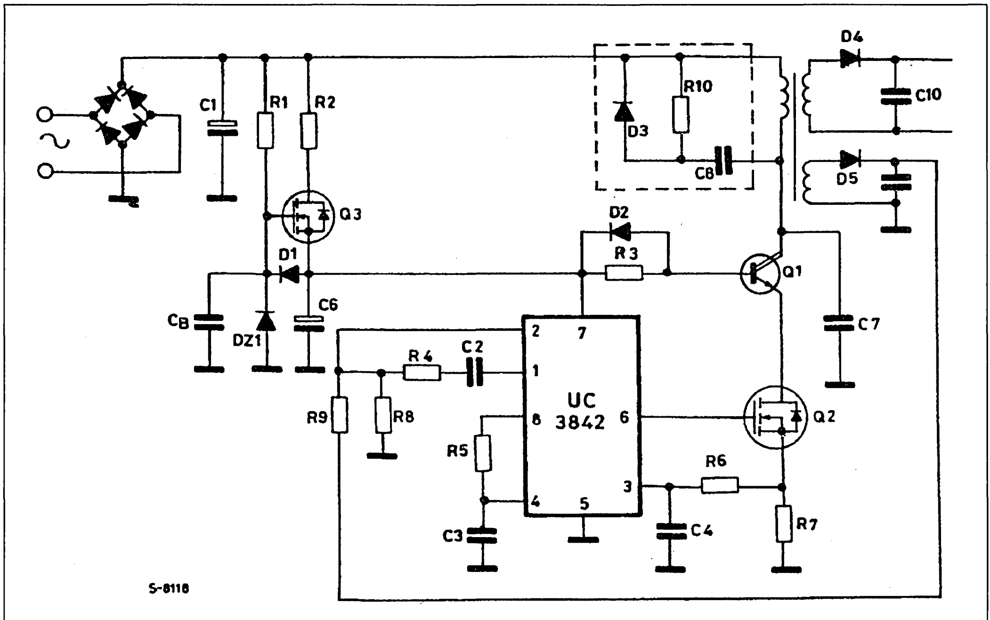
$R10 = 1/4fC8$
$P\ (R_{10}) = 1/2\ L_d\ I_D^2 \cdot f$
where :
f = switching frequency
L_d = stray inductance of the transformer
V_{OS} = maximum voltage overshoot allowed

D3 : A 400V fast recovery diode

C7 : The use of a capacitor reduces the crossover of the Darlington (3 to 6nF)

It is important to note that, the power transistor Q3 acts only at the turn-on of the power supply and when the capacitor C6 supplies more energy to the base of the Darlington and to the supply input of the IC than is returned to C6 during the turn-off of the Darlington, Q1.

Figure 6 : "Emitter Switching" Circuit.



CONCLUSION

The "emitter drive" configuration exhibts some clear differences with respect to the usual "base drive" configuration, and they can be particularly useful in switching power supply applications :

- Substantial reduction of the storage time and improvement of the fall time.

Switching frequencies of 50kHz and higher are possible

- The dynamic drive circuitry is simplified. The negative voltage supply is not required to remove the stored charge from the base. The energy needed to drive the gate of the POWER MOS is very low (180nJ per cycle).
- Extremely high ruggedness at the turn-off of the inductive load (i.e. very large RBSOA) if the high voltage bipolar part is a transistor.
- Higher power dissipation in the on-stage, due to the additional losses in the POWER MOS ($I_D^2 \cdot R_{DS\,(on)} \cdot t_{on}$).

This last point is the only disadvantage, but it is more than compensated for if switching at high frequencies. The lower switching losses (a saving each cycle) can justify the higher on-state losses (a fixed expenditure) as soon as the switching frequency is high enough, which is often the case in switching power supplies.