Proportional Regenerative Base Driver Circuit with Negative Off-State Voltage for SiC Bipolar Junction Transistors

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Abstract

A proportional base driver circuit for SiC bipolar junction transistors is presented. A current transformer feeds a proportion of the transistor’s collector current into its base. The influence of the current transformer’s magnetising current on its output current is addressed, as are minimum reset time restrictions. The circuit applies a negative off-state base-emitter voltage. Combining this and a regenerative proportional base drive current does not incur undue complexity or cost. The circuit is evaluated in a buck converter supplied from a 600V rail, switching at 50kHz and at a duty factor of 85%, and outputting 2.52kW.

1. Introduction

Desirable properties of the SiC bipolar junction transistor (BJT) include, for example, low switching and conduction losses, and a good short-circuit withstand time [1]-[9]. However, unlike MOS-gated devices, it needs a steady-state base current, \( i_B \), to hold it on as well as transient base currents for rapid switching. Its DC gain, \( h_{FE} \), is given by the ratio of collector current, \( i_C \), to \( i_B \):

\[
h_{FE} = \frac{i_C}{i_B}
\]

The base-emitter forward voltage drop (\( V_{BE} \)) of the SiC BJT is typically 3V, considerably higher than that of a silicon BJT. The \( h_{FE} \) of the SiC BJT, typically 40-100 at 25°C, is higher than that of the silicon power BJT but is not sufficiently high for the base drive current and the associated power dissipation in the power device and driver circuitry to be regarded as negligible. It is therefore desirable to supply only the \( i_B \) needed to hold the device on for a given \( i_C \); since setting \( i_B \) for the worst-case \( i_C \) incurs unwanted power losses. Also, \( i_B \) should ideally be supplied from a non-dissipative source to further improve efficiency. A regenerative circuit using a current transformer (CT) to apply a regenerative proportional base-drive current to a SiC BJT is presented in this paper. [8] addresses the challenge of efficiently providing both high transient \( i_B \) currents and a steady \( i_B \) during on-periods. Proportional drive schemes are proposed in [10]-[13] and there is work on this topic for silicon BJTs [14]-[15]. Fig. 1 outlines a regenerative base driver scheme based on a CT. After turn-on is initiated by an externally-applied current, the CT sources a proportion of \( i_C \) into the transistor’s base. Ideally, the CT’s primary current (\( i_p \)) and secondary current (\( i_s \)) are related by:

\[
\frac{i_s}{i_p} = \frac{N_1}{N_2}
\]

where \( N_1 \) and \( N_2 \) are the primary and secondary turns numbers respectively. However, due to the magnetising current drawn by the CT, the current ratio deviates from that in (2). A further consideration is that the CT’s core flux must be periodically reset to avoid saturation. In Fig. 1 \( i_s \) is applied via a diode, Dr, to enable CT core reset when TR1 is off.

Fig. 1. Outline of regenerative base-driver scheme using a CT.

The driver circuit in Fig. 1 performs three functions. Firstly, it provides the high transient base currents needed to rapidly switch TR1. Secondly, when TR1 is on, it sources a small current, \( i_{comp} \), to compensate for the current, \( i_{m2} \), drawn by the CT’s magnetising inductance (\( L_{m2} \), when referred to the secondary side). Thirdly, when TR1 is off, it provides a low-impedance path between TR1’s base and the off-state voltage level. This also diverts any current flowing out of \( N_2 \) during \( T_{off} \) away from the base, preventing unwanted turn-on. Apart from the high \( V_{BE} \) drop of the SiC BJT incurring greater losses in the device for a given \( i_B \) than in a silicon device, the higher voltage across \( N_2 \) during \( T_{on} \) causes a larger \( i_{m2} \) to be drawn by \( L_{m2} \). Dr has a forward voltage drop, \( V_r \), of approximately 1V so that the total voltage is 4V. If \( \delta \) is the duty factor at which TR1 conducts and \( T \) is the period of its switching cycle then (referred to the secondary side) a magnetising current, \( i_{m2(end)} \), given by:

\[
L_{m2} \frac{(V_{BE} + V_r) \delta T}{i_{m2(end)}}
\]

is reached at the end of the current pulse. \( i_{m2} \) can be increased by increasing the core’s effective area but this increases its bulk. If a ferrite core material is used, a high-permeability grade can be selected. However, these materials have high losses and a low Curie temperature (\( T_c \)). The latter feature can be undesirable in SiC applications as the CT is normally in proximity to the power device and a low \( T_c \) would limit the use...
of the device in high temperature locations. High-permeability nano-crystalline materials have been used in this application [13], but cost can be high.

2. Pulse current transformer operation

Pulse CT operation was investigated in [16] where, as in Fig. 1, a diode, Dr, rectifies the CT’s output signal. Apart from the changed polarities of the connections made to N2 and Dr, Fig. 2 shows the circuit in [16]. Discrete reset circuitry in the form of a voltage clamp across N2 is not used. Ceq is the lumped interturn and turn-core capacitances of the CT, and the junction capacitance of Dr. Dr’s forward voltage drop was taken as the only load driven by the CT as the output signal is fed to an operational amplifier circuit that acts as a current-to-voltage converter and presents a virtual zero-impedance load to the CT.

Fig. 2. Current pulse transformer circuit.

Three operating modes were identified in [16]. These were the discontinuous magnetising current mode (DMCM), the continuous magnetising current mode (CMCM) and the discontinuous secondary current mode (DSCM). Fig. 3 shows a rectangular current pulse, ip, with a peak value Ip applied to a CT’s primary winding. ip’s period is T and it has a duty factor \( \delta \). The centre waveforms show DMCM operation. The bottom waveforms show CMCM operation. The ripple component in ip is neglected. Vf1 is Dr’s forward voltage drop during the current pulse and Vf2 is the voltage drop seen when the core flux has reversed and i2 decays. Vf1 is invariably greater than Vf2 as, during Toff, the current through Dr is i2, but during Ton only i2 flows. As i2 is typically less than i2 by a factor of ten or more, Dr operates at a significantly different point on its v–i curve. In Fig. 3 the ratio of Vf1 to Vf2 is exaggerated to be greater than expected in a practical current-sensing circuit.

Fig. 3. CT waveforms. Top: primary current. Centre: DMCM operation. Bottom: CMCM operation.

Whereas Vf1 is the external voltage drop, in practice the voltage across Lm2 is slightly greater as it is the sum of Vf1 and the voltage developed across N2’s winding resistance by i2. Fig. 3 is simplified in Fig. 4 by neglecting Tr, where it is much smaller than T. For the DMCM mode, \( I_{m2(\text{end})} \) is given by (4). For the CMCM mode, if \( I_{m2(0)} \) is the initial current, then \( I_{m2(\text{end})} \) is found from:

\[
\omega_r = \frac{1}{\sqrt{L_{m2}C_{eq}}} \tag{6}
\]

The damping effect of core losses is neglected in (5) and (6). When the resonant half cycle has elapsed after a time, Ton, \( I_{m2} \) then flows through Dr where its Vf causes \( I_{m2} \) to decay to zero over a period denoted as \( T_\text{e} \). Some of the \( I_{m2} \) product lost during Ton is thus returned to the CT’s output terminals during Toff. Advantageously in a sensing application, average current droop can therefore be less than that inferred from the peak current drop in (4). In the DMCM mode \( I_{m2} \) decays to zero before the current pulse is re-applied but in the CMCM shown in the bottom waveforms \( I_{m2} \) has not decayed to zero at this point. Although \( I_{m2} \) passes through zero in the CMCM mode, the term continuous is used here to refer to a state where \( I_{m2} \) is only at zero instantaneously. Vf1 is Dr’s forward voltage drop during the current pulse and Vf2 is the voltage drop seen when the core flux has reversed and \( I_{m2} \) decays. Vf1 is invariably greater than Vf2 as, during Toff, the current through Dr is i2, but during Ton only i2 flows. As i2 is typically less than i2 by a factor of ten or more, Dr operates at a significantly different point on its v–i curve. In Fig. 3 the ratio of Vf1 to Vf2 is exaggerated to be greater than expected in a practical current-sensing circuit.
\[ I_{m2(\text{end})} = I_{m2(0)} + \frac{V_{f1} \delta T}{L_{m2}} \]  
\[ I_{m2(0)} = -I_{m2(\text{end})} + \frac{V_{f2}(1 - \delta) T}{L_{m2}} \]

Combining (7) and (8) yields:
\[ I_{m2(\text{end})} = \frac{T}{2L_{m2}} \left[ V_{f2}(1 - \delta) + V_{f1} \delta \right]. \]  

If \( V_{f2} \) is close in value to \( V_{f1} \) then:
\[ \delta_{TH} = \frac{T_{on}}{2L_{m2}} \]  
If \( V_{f2} \) is much smaller than \( V_{f1} \) then:
\[ \delta_{TH} = \frac{\delta T V_{f1}}{2L_{m2}}. \]

The threshold duty factor, \( \delta_{TH} \), at which the transition between the DMCM and CMCM modes takes place is given by:
\[ \delta_{TH} = \frac{T_{on}}{2L_{m2}}. \]

In summary, with ideal resonant resetting and \( V_{f2} \leq V_{f1}, I_{m2(\text{end})} \) is always restricted to the value in (10), even at duty factors approaching 100% in the CMCM mode, the initial direction of \( I_{m2} \) when \( I_{p} \) is applied is such that \( I_{2} \) exceeds the value given by (2). That is, at the beginning of the current pulse, there is an initial oversupply of \( I_{2} \) as \( I_{m2} \) has reversed. The ratio of \( V_{f2} \) to \( V_{f1} \) is given by:
\[ V_{f2} = k V_{f1}. \]

where \( k \) is a constant. Fig. 5 shows \( I_{m2(\text{end})} \) against \( \delta \) for different values of \( k \) for the idealised circuit operation in Fig. 4, where losses in the CT are neglected and \( T \) is taken as much greater than \( T_{on} \). Quantities in Fig. 5 are normalised to one. The curves for \( k = 1 \) and \( k = 0 \) in the CMCM mode represent the situations in (10) and (11) respectively. Curves for intermediate values of \( k \) (0.25, 0.5 and 0.75) are also shown. If the series combination of \( V_{f} \) and \( V_{H} \) is close to an ideal voltage sink then \( I_{m2} \) is independent of \( I_{2} \). In the extreme case, all of \( I_{2} \) disappears into \( I_{m2} \). This is the DSCM mode of operation described in [16]. As mentioned in Section 1, where there is a need to cater for very low currents, \( I_{2} \) has to be augmented with a small current, \( I_{comp} \).
3. Base driver circuit with regenerative feedback and negative off-state voltage

3.1. Proposed topology

Fig. 6 shows the proposed topology. The emitter follower pair, TR2 and TR3, sources a current pulse into the base of TR1 to start conduction and sinks a current pulse to turn it off. These pulses are supplied via Cp. As in, for example, [8] a resistor, R2, is included to suppress oscillations. Whilst Cp is used here to provide high transient currents for rapid switching, it is noted that an alternative technique for realising this is to use a current source driver [17]. R1 provides a current, $i_{\text{comp}}$, to cover the shortfall in $i_{B}$ due to the $i_{au2}$ drawn at the end of a current pulse. A negative off-state voltage is often needed from gate or base driver circuitry to avoid unwanted turn-on due to externally-applied $dV/dt$ across the power electrodes injecting currents through Miller capacitances.

$$V_{B} = V_{T1} = 4V.$$  

The proposed circuit was used to drive a GA06JT12-247 SiC BJT [18] (TR1) in a buck converter, Fig. 7, with a SDP30S120 SiC diode (D1). No forced cooling was applied. With respect to Fig. 6, $V_{\text{rail}} = 600V$ and $I_{\text{load}} = 4.95A$, TR1 switches at $f_{\text{sw}} = 50kHz$ and $\delta = 85\%$. The output power is approximately 2.52kW. For experimentation, a large output choke inductance was used to give a peak ripple content in $I_{\text{load}}$ of approximately 2% of its average. Local decoupling of $V_{\text{rail}}$ was provided by two 5.0µF, 800V polypropylene capacitors in parallel. The base driver supply rails were $V_{\text{HRI}} = 12V$ and $V_{\text{BRI}} = -5V$. A TN13/7.5/5 core in 3C90 material was chosen for CT1. Its data are: $A_{e} = 12.2mm^{2}$, $l_{e} = 30.1mm$, $V_{r} = 368mm^{3}$, $\mu_{e} = 2300$ and $T_{e} \geq 220^\circ C$ [19]. $h_{FE}$ is estimated at 40 for the GA06JT12-247 device when allowance is made for operation at an elevated temperature. Applying a factor of safety of 1.5 yields a required current gain of 26.7. This gives $i_{b0} = 185mA$ for $i_{c0} = 4.95A$. $N_{1}$ was formed from two turns. $N_{2}$ was formed from 52 turns of 0.315-mm diameter polyurethane-insulated copper wire to give a turns-ratio close to 26.7. Applying the data from [19] and $N_{2} = 52$ yields $L_{m2} = 3.17mH$. Normally a CT only carries one primary turn, but two were used here as this increases $L_{m}$. The turns-ratio was set to yield the nominal current gain given by (2), and alternative turns-ratios were not considered.

$$f_{\text{sw}} = 50kHz \text{ gives } T = 20\mu s.$$  

As mentioned in Section 3.1, $V_{B}$ is taken as 3V and $V_{f}$ is taken as 1V. Putting these values and that of $L_{au2}$ into (10) gives an expected current shortfall of 13mA at the end of a pulse. When setting $i_{\text{comp}}$, a 50% safety margin was again applied, and it was set at 20mA. Using (10) assumes that resonant CT operation is effective in restricting the maximum droop to less than that calculated in the normal way and that core losses may be neglected. Apart from CT1, component data are: $TR2 = ZTX651$, $TR3 = ZTX751$, $TR4 = FDT458P$, $D2 = D3 = D4 = ES1A-13-F$, $R1 = 260\Omega$, $R2 = 2.2\Omega$, $Cp = 22nF$, $D$ = 1DD03SG60C. The voltage across $R1$ during $T_{on}$ is $V_{\text{DRI}}$, $V_{\text{CTR1}}$, $V_{\text{BE}}$ where $V_{\text{CTR1}}$ is the on-state voltage drop across TR2, approximated at 1V. As $V_{\text{DRI}}$ is 12V, 8V is therefore applied across $R1$ and it was made 360Ω to give $i_{\text{comp}} > 20mA$. $Cp$ and $R2$ were set experimentally. U1 and U2

Fig. 6. Proposed circuit driving SiC BJT in a buck converter.

TR4 is a P-channel MOSFET. When TR1 is off, TR4’s gate is held high. It is therefore off and disconnects the secondary winding of CT1 from TR1’s emitter. This allows a negative $V_{BI}$ to be applied without drawing current through $N_{2}$. $V_{BI}$ is given by the sum of the forward voltage drops of $03$, $Dr$, $02$ and the voltage drop across $N_{2}$. Each of these is approximated as 1V, therefore giving $V_{B} = V_{f} = 4V$ and, from (13), $k = 1$.

Fig. 7. Experimental circuit driving SiC BJT in a buck converter. TR1 and D1 are mounted on the heatsink underneath the PCB.

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Fig. 7. Experimental circuit driving SiC BJT in a buck converter. TR1 and D1 are mounted on the heatsink underneath the PCB.
are 4000-series Schmitt gates. Dr is a SiC Schottky diode and a beneficial feature of using a SiC diode here is its combination of a high reverse voltage rating and low junction capacitance for a given current rating. Its $V_f$ is high compared to other diode types, but, given the SiC BJT’s high $V_{BE}$, the extra voltage drop is a small proportion of the total voltage across $N_2$ during $T_{on}$. With $k = 1$, the peak flux density excursion, $B_{peak}$, in the CT’s core material is given by:

$$B_{peak} = \frac{TV_1}{2N_2A_e}$$

Putting values into (14) yields $B_{peak} = 66mT$. At $f_{sw} = 50kHz$ and $B_{peak} = 66mT$, core losses are estimated at less than 4mW from the graphical data in [19]. Whilst this data is for sinusoidal flux density excursions, and is given for 100°C, the impact of core losses on overall efficiency is nonetheless considered sufficiently small to be neglected.

3.3. Experimental results

3.3.1. Waveforms

Fig. 8 shows switching-frequency waveforms. Figs. 9 and 10 show waveforms at TR1 turn-on and turn-off respectively. The “OV” point used for measuring voltages in this section is at the emitter of TR1. As one end of $N_2$ is connected to $-V_{DC}$ during $T_{off}$, it is seen that a negative base voltage is superimposed onto $v_2$. The over-voltage in $v_{CE}$ at TR1 turn-off was measured at around 6V, 1% of the supply voltage.

![Fig. 8. Switching-frequency waveforms showing a full 50kHz switching cycle. $v_{BE} = 10V/div$, $i_{LOAD} = 10A/div$, $v_2 = 50V/div$, $v_{CE} = 500V/div$. Time scale = 2µs/div.](image1)

![Fig. 9. Waveforms at TR1 turn-on. $v_{BE} = 10V/div$, $i_2 = 200mA/div$, $v_{CE} = 500V/div$. Time scale = 50ns/div.](image2)

![Fig. 10. Waveforms at TR1 turn-off. $v_{BE} = 10V/div$, $i_2 = 200mA/div$, $v_{CE} = 500V/div$. Time scale = 50ns/div.](image3)

3.3.2. Losses

The heatsink steady-state temperature rise above ambient, $\Delta T$, was 32.9°C. Its thermal resistance, $R_h$, was measured at 1.77°C/W with a prior thermal superposition test. A loss of 18.6W was therefore estimated. At 2.52kW output power the estimated efficiency in the semiconductor devices is therefore 99.3%. The currents drawn from the driver circuit’s 12V and 5V supply rails were measured at 32mA and 3mA respectively, giving a power consumption of 399mW. This represents approximately 2% of the total loss.

4. Conclusions and discussion

A simple SiC BJT base driver circuit with regenerative proportional feedback using a CT and a negative off-state voltage has been presented. These features are combined without incurring undue complexity or cost. The measured power of 399mW drawn covers both the power associated with switching and with supplying $i_{comp}$. The power needed to provide the entire steady-state $i_B$ of 185mA at $\delta = 85%$ from a 12V rail would be 1.887W (not including losses incurred through switching the BJT). Even if this power were derived from a lower-voltage rail at, say, 5V then 786mW would be drawn. Some further points are raised:

- The on-state driver power is often supplied from an SMPS to give isolation. The SMPS incurs losses. Assigning a typical efficiency of 80% to this stage would give a power draw of 983mW if the SMPS provides a 5V rail.
- With CT feedback, the loss in Dr is accounted for by taking $V_f = 1V$, and this gives a loss of 157mW when conducting 185mA at $\delta = 85%$.
- If a lower-voltage rail is used to supply the on-state $i_B$ with reduced losses, the driver circuitry normally needs a total of three supply rails: one to give a negative off-state voltage, one to supply the steady-state base current and one at a higher voltage to enable rapid switching.
- If an SMPS with isolation is used to supply the driver power requirements for a SiC BJT, this is a relatively costly way of supplying power due to the isolation requirement. (Whilst CTs can provide isolation, this is not needed here.)
- When setting $i_{comp}$, it was assumed that the CT incurs no losses and that the curves in Fig. 5 are appropriate. On the other hand, the value of $k = 1$ applied is pessimistic.
Although the effect of core losses on efficiency is small, a challenge when designing this circuit is in gauging the effect of the losses on the magnetising current left at the end of a resonant resetting action.

- With regenerative feedback, a further reduction in losses in both the driver stage and the power device is attainable due to the reduced $i_c$ supplied when operating with $i_c$ below its maximum value. This is particularly advantageous where the power converter operates with high ratios of maximum to average $i_c$, for example, in some machine-drive applications.
- A high duty factor is attainable, but duty factors of 100% are desirable in some converters. However, expected applications include DC-DC converters where duty factors of 100% are not generally required. A negative off-state base voltage is desirable in bidirectional DC-DC converters using bridge-legs [2] where the power devices are subjected to high $dV/dt$. Other applications include half-bridge and full-bridge SMPS circuits where duty factors are less than 50% and the devices are also subjected to high $dV/dt$. (In these cases a reduced CT reset voltage can be used as longer reset times are available.)
- A CT in series with a power device is expected to introduce parasitic inductances but the overshoot voltage observed across the power device at turn-off is modest, at around 1% when switching 4.9SA from a rail voltage of 600V over approximately 50ns.
- A high-voltage rectifier diode with a low junction capacitance and no other reset circuitry is desirable to allow CT operation at high duty factors. 90% duty factor can be attained at 50kHz.
- The current, $i_{comp}$, supplied to compensate for the loss of current into the CT’s magnetising branch has a simple rectangular profile. The use of ramp profiles to further reduce losses has not been investigated.
- Some aspects of the circuitry have not been optimised. Optimisation of the timings between the main drive signal output from TR2 and TR3, and the signal applied to TR4 in Fig.6 has not been investigated. The speed-up components, R2 and C, and the rail voltages have not been optimised. The construction of the CT has not been optimised to reduce stray capacitance, a factor that determines the maximum duty factor. An aggregate rail voltage of 17V was used due to the limited voltage rating of the 4000-series logic gates. A buffer stage with a higher voltage capability would be expected to reduce switching losses.

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