SGS-THOMSON MICROELECTRONICS

APPLICATION NOTE

BIPOLAR JUNCTION TRANSISTORS, POWER MOSFETs OR IGBTs IN RESONANT CONVERTERS

by P. Fichera

ABSTRACT

Resonant switch topologies operating on the principle of zero-current and zero-voltage switching are discussed. Their advantages with respect to the conventional PWM converter are shown.
The main advantages and disadvantages of some resonant structures are considered when the ideal active switch is either a bipolar junction transistor (BJT), a power MOSFET, or an insulated gate bipolar transistor (IGBT). Practical examples of power semiconductor choices in these resonant topologies are given.

1. INTRODUCTION

The main advantage of resonant structures is the reduction of switching losses, and also less stress on the electronic switch occurs compared with PWM structures.
These two advantages (lower switching losses, less device stress) are generally paid for in terms of higher conduction losses.
The purpose of this paper is to present the elements that permit the most suitable choice of power semiconductor to be made in some types of resonant structures.

2. RESONANT SWITCH

The resonant switch represents the basic element of converters with resonance or quasi resonance. It consists of semiconductors and LC resonant elements. Depending on the configuration of the resonant elements one can obtain the structures of fig. 2 [1,2]
A family of quasi resonant converters (QRC) is obtained by simply replacing the conventional chopper power switch with a resonant switch as shown in fig. 3.
In each case there is controlled switching either at turn-on in the ZCS or at turn-off in the ZVS.

3. ZERO-CURRENT AND ZERO-VOLTAGE SWITCHED QUASI-RESONANT CONVERTERS (ZCS-QRC and ZVS-QRC)

3.1 ZCS-QRC

This configuration permits the switch to be controlled at turn-on, while the turn-off occurs with zero current. The switched current, I_S is a quasi-sine-wave and is reduced to zero at turn-off.
Switching losses are due to:
1) turn-on switching (although these losses are much lower in comparison to the PWM converter),
2) internal discharge of the power semiconductor output capacitance C_{oss} at turn-on (see waveforms fig.4)[7]
3) rectifier diode switch-off.

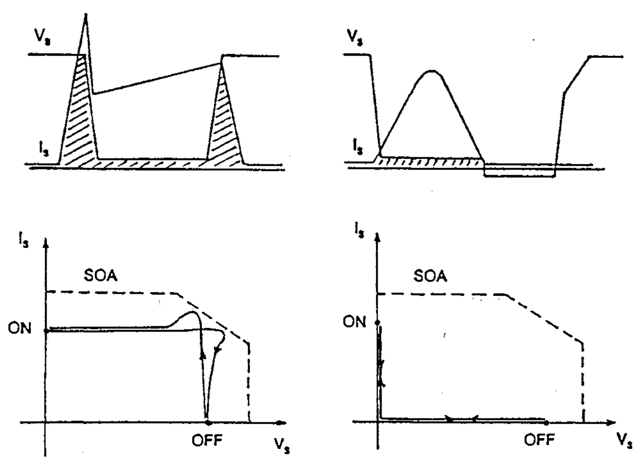


Fig. 1 Hard switching Soft switching

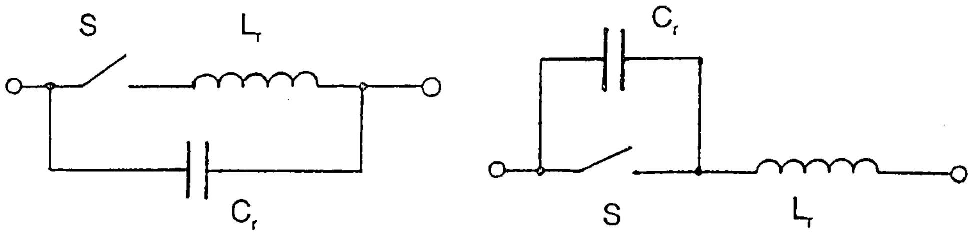


Fig.2 ZC-QR switch (thyristor) ZV-QR switch (Dual thyristor)

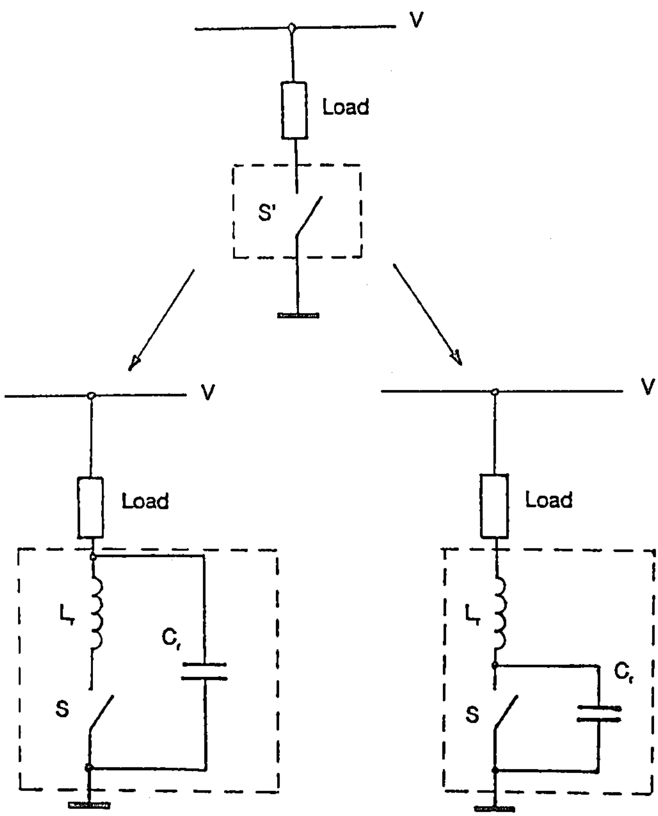


Fig. 3 ZCS-QRC ZVS-QRC

3.2 ZVS-QRC

This configuration permits the switch turn-off to be controlled; the turn-on occurs with zero voltage (the capacitive turn-on problem of ZCS-QRC is eliminated). This operating mode is called "dual thyristor" because the switching properties follow the duality rules of the thyristor, i.e. spontaneous turn-on at zero voltage and controlled turn-off [Ref.1].
The voltage across the switching device is quasi sine wave and is reduced to zero at turn-on. The only switching losses occur at turn-off as shown in the waveforms of fig.5.

4. WHICH SWITCH?

We are interested in analysing the behaviour of the two quasi resonant (QR) topologies when the active switch, S, is a BJT, a Power MOSFET or an IGBT.
Table 1 shows the main features of each power semiconductor (V > 400V).
Because the cost of a power semiconductor varies with the chip-size the curve of fig.6 gives a cost analysis.

4.1 THE PHYSICAL LIMIT OF A POWER BIPOLAR TRANSISTOR.

The presence of storage time, t_s, limits the maximum frequency of operation. Extra losses are due to the presence of $V_{CEsat(dyn)}$ at turn-on.

4.1.1 A BIPOLAR JUNCTION TRANSISTOR IN ZCS.

The storage time, t_s, limits the duration of t_d between the time that the current is zero and the re-applied voltage (t_d is in the order of t_s). The transistor voltage does not reach zero instantaneously at turn-on ($V_{CEsat(dyn)}$). This phenomenon could increase the conduction losses for a short time.

4.1.2 A BIPOLAR JUNCTION TRANSISTOR IN ZVS.

The storage time, t_s, limits the maximum frequency of operation of the converter.

4.2 THE PHYSICAL LIMIT OF A POWER MOSFET.

If there are no economic limits the Power MOSFET is almost a perfect component for many applications, but has two real limitations:
a) internal capacitance;
b) the presence of the body drain diode. This diode is:
b.1) a good diode during its turn-on behaviour (very low peak voltage)
b.2) a poor diode during turn-off.

4.2.1. A Power MOSFET in ZCS

The Power MOSFET is not adapted well to ZCS operation because the body drain diode is subjected to very hard stress due to the recovery phenomenon.
The only solution when using a Power MOSFET in the ZCS mode at high frequency (> 20kHz) is to add a fast diode (D2) and a Schottky diode (D1) as shown in fig.8
The maximum frequency will be limited by the losses due to the Power MOSFET internal capacitance.

4.2.2. A Power MOSFET in ZVS.

In this type of operation, as in the "dual thyristor" operation, the body drain diode is not subjected to stress after it's current becomes zero.
In addition, the turn-on occurs at zero voltage and consequently the energy stored in the parasitic MOS capacitance is not lost. The nature of a Power MOSFET makes it very suitable for ZVS operation.

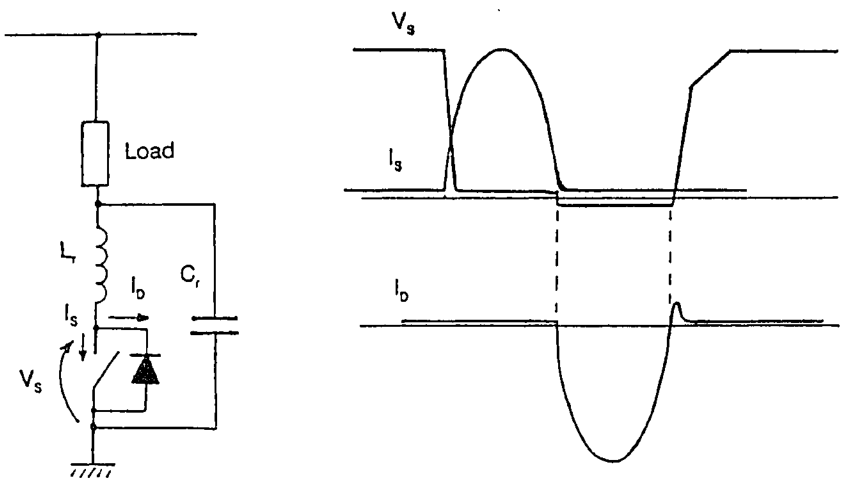


Fig. 4 ZCS-QRC (with bidirectional current switch) and corresponding waveforms.

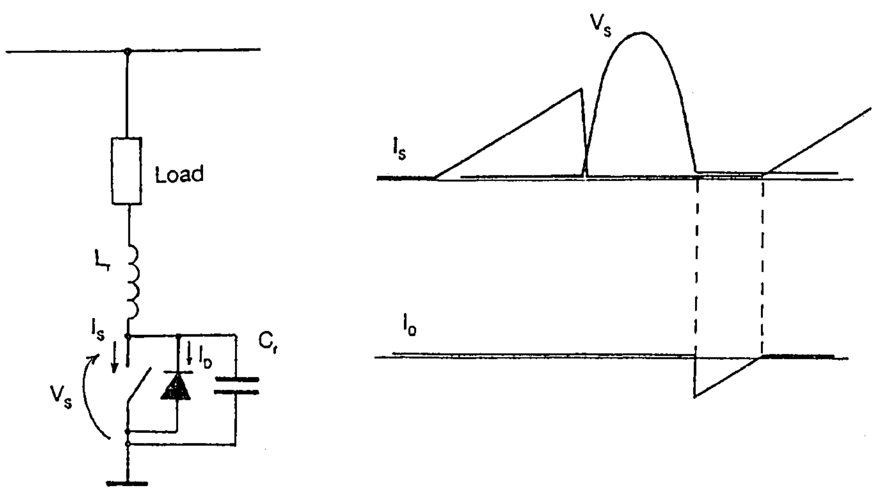
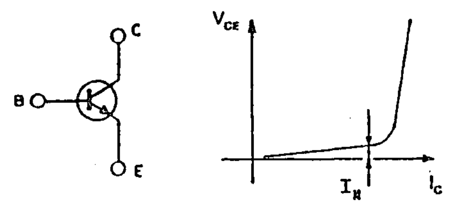
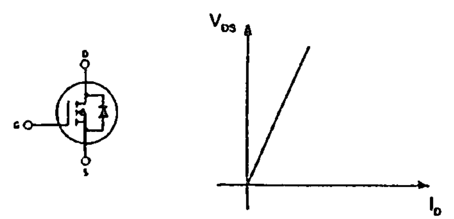
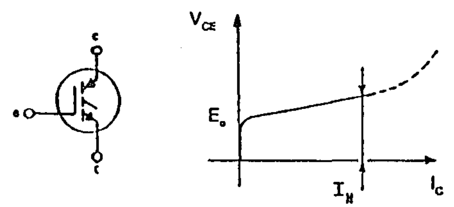


Fig. 5 ZVS-QRC (with unidirectional voltage switch) and corresponding waveforms.

	**ADVANTAGES** - Very low on-state voltage drop - Reduced silicon area - Low cost	**DISADVANTAGES** - Cost of base drive - Storage time (t_s)
	**ADVANTAGES** - Low voltage drop is obtained with large silicon area - Cheap gate drive - Reduced turn-off delay time - Anti-parallel body-drain diode can be used	**DISADVANTAGES** - Cost - In some cases dangerous to use body drain diode
	**ADVANTAGES** - Low voltage drop at high current density - Cheap gate drive	**DISADVANTAGES** - Presence of threshold voltage E_o - Current tail effect at turn off

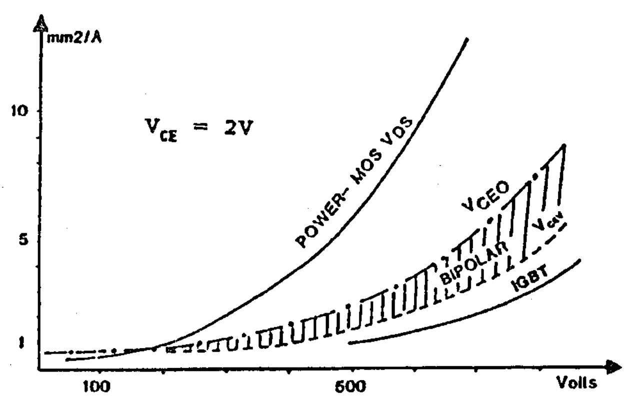
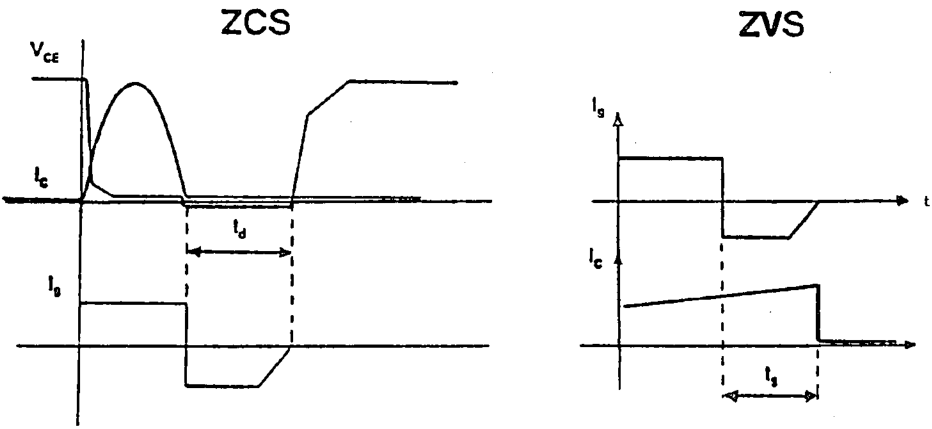


Fig. 6 Chip surface vs maximum rated voltage



- Presence of a dead time t_d which limits f_{max} (t_d is in the order of ts and depends on the base drive circuit)
- $V_{CEsat\ (dyn)}$

- The storage time (t_d) limits f_{max}

Fig. 7 - The physical limit of Power Bipolar Transistors in quasi resonant circuits.

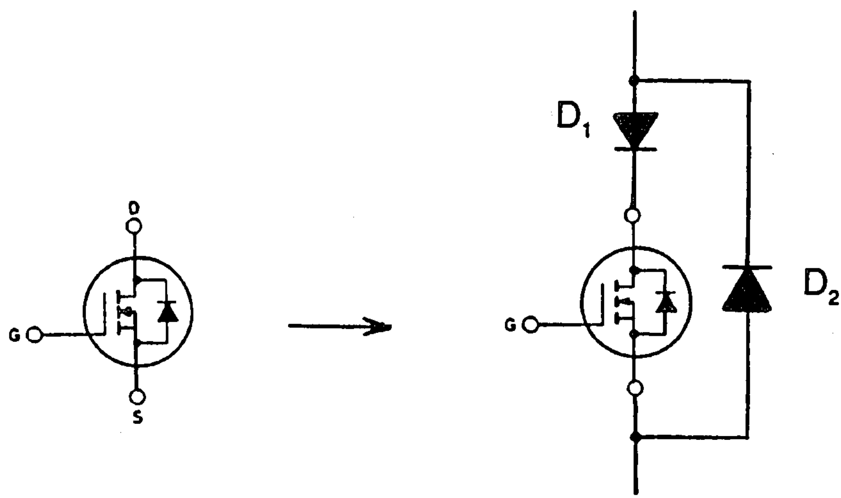
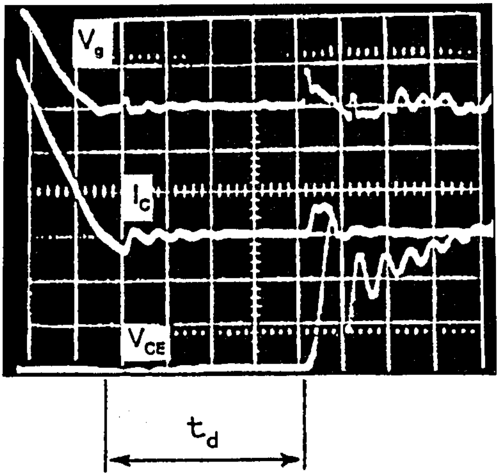


Fig.8 Disabling a Power MOS body drain diode



- $t = 0.2$ ms/div.
- $V_{CE} = 100$V/div.
- $I_c = 5$A/div.
- $V_g = 5$A/div.
- $T_c = 50°C$
- $dV/dT = 400$V/ms

Fig. 9 IGBT behaviour during turn-off in ZCS mode. The waiting time, t_d, is not enough and a current peak, bigger than that due to the capacitive effect, appears due to the effect of the re-applied voltage.

4.3 THE PHYSICAL LIMIT OF AN IGBT.

The IGBT has an economic advantage in comparison to the Power MOSFET (smaller silicon area for the same voltage drop), but its turn-off is influenced by the bipolar transistor section. At turn-off, the minority carriers which remain in the base-collector junction of the bipolar section increase the turn-off losses (current tail effect). If the IGBT turn-off is controlled by only acting on the gate, it is necessary to wait for a time, t_d, before re-applying the voltage, otherwise extra losses generated may make the device fail due to thermal run-away. This time, t_d, must be longer than the duration of the current tail effect (in the order of 3-4μs for a 1000V IGBT)[8].

4.3.1 AN IGBT IN ZCS.

The frequency is limited by t_d, consequently the IGBT in ZCS can attain a relatively high frequency (around 120kHz). The losses due to the internal IGBT capacitance in this frequency range are smaller than Power MOSFET losses. In fact, for a given current rating the IGBT capacitance is much smaller than the Power MOSFET capacitance. The only other limitations are the conduction losses.

4.3.2 AN IGBT IN ZVS.

The main limitation of the IGBT in this configuration is the high switching losses due to the current tail effect. A larger value capacitor, Cr, acting as a snubber, reduces dV/dt at turn-off and, consequently, turn-off losses [5].

SUMMARY OF THE PHYSICAL LIMITS OF SEMICONDUCTORS

TABLE II

BJT	- Storage time —> frequency limitation - $V_{CEsat(on)}$ —> turn-on losses	- Storage time —> frequency limitation
Power MOSFET	- Recovery of body drain diode - Capacitive turn-on losses at high frequency - Conduction losses	- High voltage
IGBT	- Deadtime —> frequency limitation - $V_{on\text{-}state}$ —> conduction losses	- Turn-off tail problems

5. TWO PRACTICAL EXAMPLES.

5.1 The first example we are going to consider is a single-switch ZCS-QRC with the following characteristics:
1.5kW - 50kHz - 220V mains.
The switch current waveform will be characterized by:

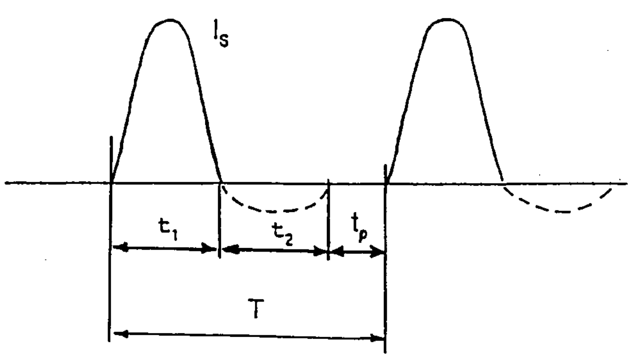
Because the time $t_2 < t_1 = 8\mu s$ (conduction time of the antiparallel diode), this time is large enough if compared with the fall-time, t_f of an IGBT (< $1\mu s$) or the storage time, t_s, of a bipolar transistor ($1\text{-}3\mu s$) there are no problems in using these semiconductors.

$t_1/T = 0.4$

$I_{AVG} = 5A$

$I_{peak} = 20A$

$I_{RMS} = 9A$



ANALYSIS OF DIFFERENT POWER SEMICONDUCTORS

(Devices selected with similar current rating)

BUF420	450/850V - 20A ETD Bipolar Junction Transistor) [54 square mm. silicon area]. Conduction losses calculation gives P_{cond} = 8.1W.
TSD4M451	(450V/$R_{DS(on)}$ = 0.15 Ohms at 100°C ISOFET) [176 square mm. silicon area] To obtain conduction losses comparable to BJT losses, it is necessary to choose a large silicon area Power MOSFET (P_{cond} =12.1W). The main disadvantages are: 1) turn-on capacitive losses 5.5W at 50 kHz due to the big chip-size (176 square mm total silicon area) 2) extra conduction losses (3W) due to the use of a power Schottky diode
STGH20N50	IGBT 500V/20A) [32 square mm. silicon area] conduction losses calculation[5] gives P_{cond} = 12W.
STGP10N50	(IGBT 500V/10) [16 square mm. silicon area]. Due to its over current capability a 10A/500V IGBT can be used in this application. Conduction losses calculation[5] gives P_{cond} = 18.6W.

CONCLUSION: IGBTs and BJTs are concurrent solutions. IGBT advantages are:

1) gate drive simplicity
2) saving in silicon area.

BJT advantage is due to the optimum ratio of conduction losses to device cost. However the storage time limits the BJT frequency operation to no more than 70kHz.

5.2 The second example we are going to consider is a single - switch ZVS-QRC with the following characteristics:

300W - 150kHz - 110V mains (± 20%).

The switch current waveform will be characterized by:

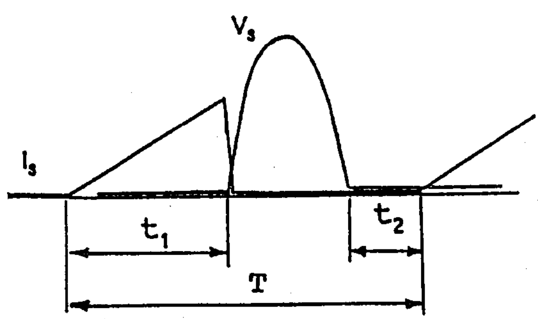
$(t_1+t_2)/T = 0.6$

I_{AVG} = 2A

I_{peak} = 10A

I_{RMS} = 3.65A

$V_{s(peak)}$ = 750V (max.)



ANALYSIS OF DIFFERENT POWER SEMICONDUCTORS.

(Devices selected with similar current rating)

BUF410 — (450/850V - 10A ETD bipolar junction transistor). The storage time t_s (1 - 3µs) is comparable with $t_1 + t_2 = 4\mu s$. The use of a BJT at this frequency of operation is not possible. For the ETD BJT technology an upper frequency of operation is around 100kHz.

STH9N80 — (800V/$R_{DS(on)}$ = 1.5 Ohms at 100°C Power MOSFET). Conduction loss calculations give P_{cond} = 20W. Turn-off losses are negligible. The Power MOSFET body drain diode is not subjected to any stress and can be used as an anti-parallel diode.

STGH8N100 — (1000V/8A IGBT). At I_{switch} = 10A and dV/dt = 1000V/µs the IGBT turn-off losses can be evaluated at around 0.6-0.7mJ/Hz. It means roughly 105W at 150kHz only for turn-off losses. Conduction loss evaluations give P_{cond} = 4.6W.

CONCLUSION: For this single-switch application, the Power MOSFET is the most suitable choice.

At the given frequency of operation and duty cycle one obtains: $t_1 + t_2 = 4\ \mu s$

SUMMARY OF EXAMPLES

The physical limit of the semiconductors examined and the examples taken into consideration show that in a single switch QRC a correct choice of the switch is the following:

IF

ZCS - QRC:

Bipolar Junction Transistor (f < 70kHz)

Power MOS (f < 20kHz).

Higher frequency of operation could be obtained disabling internal body drain diode or using a FREDFET.

IGBT (f < 120 kHz)

IF

ZVS - QRC:

Bipolar Junction Transistor (f < 100kHz)

(f = 0.5 to 1MHz)

IGBT (f < 20 - 30kHz)

Power MOSFET (0.5 MHz -1MHz)

CONCLUSION

Each power semiconductor we took into consideration shows physical characteristics that limit its operation in single switch quasi resonant converters.

In particular the bipolar solution is acceptable for both QRC topologies at frequencies lower than 100kHz. The power MOSFET solution is disadvantageous in applications at high

RMS currents such as those in discontinuous ZCS-QRC. Despite this, power MOSFETs are most suited to the ZVS-QRC due to their fast switching times and the possibility of using the body drain diode even with their inferior performance.

The IGBT solution is today the most popular for resonant and quasi resonant applications. However in ZCS-QRC where its use seemed to permit very high working frequencies, the dynamic dV/dt phenomenon may lead the device to fail due to thermal runaway. This can be overcome by careful design and the use of an efficient heatsink.

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4) J.M.Bourgeois, B.Maurice, *"Losses of insulated gate bipolar transistor in H.F. Resonant Converters"*. IEEE Industry Applications Society Annual Meeting, 1989 - pp.1197-1204.

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7) K.Rischmuller, *"Improve efficiency of high frequency power conversion designs"* Proceedings of P.E.C. California, February 1990.

8) R.Letor, M.Melito, *"Safe behaviour of IGBTs submitted to a dV/dt"*. Power Conversion Munich June 25-28, 1990 (same proceedings).

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APPLICATION NOTE

AN ANALYSIS OF LOSSES IN AN IGBT

by P. Fichera

1. INTRODUCTION

Insulated gate bipolar transistors are now being used in a variety of switching applications. These range from automotive ignition, where they replace the mechanical contact breaker, to electric motor drives, where they provide an economic, easy to drive chopper switch with high voltage capability.

More recently work has been done in using these devices in various types of power supplies.

They are attractive to use due to the high impedance input, a MOS gate that requires a mimimum of only 8V and microjoules of energy to turn it on and off and the bipolar nature of the output that makes them capable of controlling high current densities.

To obtain the optimum performance from IGBTs it is necessary to understand the limits imposed by the structure of the device and their particular operating conditions.

This paper looks at the use of IGBTs in chopper circuits and shows how to evaluate the losses during switching and conduction.

2. LIMITING FACTORS FOR IGBTS IN CHOPPER CIRCUITS

Chopper circuits operate at frequencies determined by the nature of the application and of the power switch employed to control the current flow. As is the case with Power MOSFETs, power is dissipated in IGBTs at turn-on of the device, during conduction and at turn-off. The major difference between IGBT and Power MOSFET switching losses occurs in the turn-off switching behaviour. Figure 1 illustrates the typical losses for an IGBT used in a chopper application.

2.1 TURN-ON LOSSES

It is not sufficient to know the rise time, tr, of the turn-on current. The free-wheeling diode used in conjunction with the IGBT, figure 8, is responsible for a large amount of the losses as a result of its reverse recovery current. Within a given application it is necessary to know the $(dI_D/dt)_{on}$ for this diode in order to evaluate the reverse recovery current, I_{RM}. Once I_{RM} is known it is possible to calculate the turn-on losses.

2.2 CONDUCTION LOSSES

The following simple expression shows how to calculate the conduction losses.

$$P_{on} = E_o\, I_{AVG} + R_o\, I^2_{RMS}$$

where:

P_{on} = on-state power dissipation

I_{RMS} = RMS current value for the application

I_{AVG} = average current value of the application

E_o, R_o = are parameters defined by the IGBT output characteristic I_c, V_{ce} - see figure 2.

E_o = abscissa of the intersection between the tangent to the output characteristic calculated at $I_c = I_o$ and the V_{ce} axis.

R_o = inverse slope of the tangent to the output characteristic curve I_C, V_{CE}, calculated at $I_c=I_o$.

The area B in figure 1 illustrates these losses.

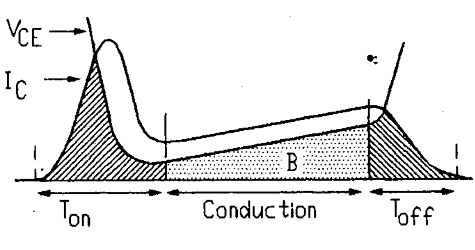


Fig. 1 - Typical IGBT losses

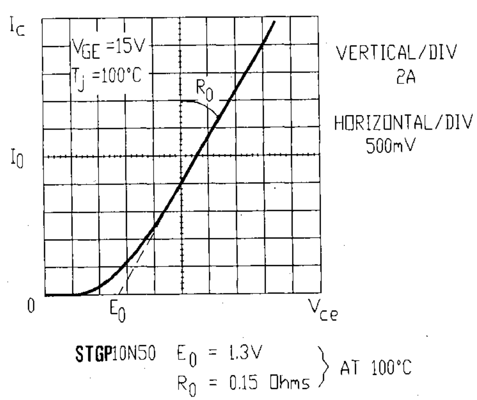


Fig. 2 - Output characteristics of STGP10N50

2.3 TURN-OFF LOSSES

Calculation of the turn-off losses in an IGBT requires more information than just the fall time, t_1. On its own it leads to erroneous results. It is necessary to know how other parameters influence these losses.
Most care has to be taken with the current tail phenomenon of the IGBT when it is operated in hard switching. Two parameters define the current tail: its amplitude, I_t, and its duration, t_t.

A) THE INFLUENCE OF THE SUPPLY VOLTAGE ON TURN-OFF LOSSES.

The supply voltage and the current tail amplitude are directly proportional. However, the duration of the tail remains almost constant when the supply voltage is varied.

B) THE INFLUENCE OF DV/DT ON THE TURN-OFF LOSSES

A low dV/dt value at turn-off (imposed by an external circuit, e.g. a snubber) reduces the current tail amplitude, I_t. The tail duration does not change when dV/dt is varied.

C) THE INFLUENCE OF TEMPERATURE ON TURN- OFF LOSSES.

Operating temperature affects the duration and amplitude of I_t and t_t. Experimental analysis shows that both increase in value by the same percentage as the temperature increases.

D) THE INFLUENCE OF THE GATE RESISTANCE, $R_{g(off)}$.

The gate resistance does not affect the current tail. Varying $R_{g(off)}$ controls the slope of dV/dt at turn off and consequently can give some reduction in the turn-off losses. A minimum value of $R_{g(off)}$ is required to prevent oscillations occuring during turn-off (as is the case with power MOSFETs).

5 CALCULATING CONDUCTION LOSSES

When calculating conduction losses at 100°C it is better to base the calculation on the output characteristics of the IGBT (I_c versus V_{ce}) at a given V_{ge}.

Table 1. *Additional parameters required to calculate turn-off losses in IGBTs.*

V	- the re-applied supply voltage
dV/dt	- slope of re-applied supply voltage
T_j	- junction temperature
$R_{g(off)}$	- gate resistance at turn off.

3. SAMPLE CALCULATIONS.

It is possible, using the curves given in figure 9 and the energy curves characteristic of figure 10 for different operating conditions, to calculate the switching losses for a given set of conditions. This in turn allows the maximum operating frequency for the IGBT to be calculated.
The basic circuit in figure 8 shows the configuration used for the STGP10N50 500V, 10A IGBT and its switching waveforms.

3.1 CALCULATION OF THE TURN-ON LOSSES AT T_j = 100°C.

The value of the gate resistor during turn-on is 47 Ohms. Using the graph in figure 9a this gives a value for dI_D/dt_{on} of 100A/µs. As the IGBT controls dI/dt it follows that the recovery current of the free-wheeling diode can be determined from the diode datasheet (graph of I_{RM} versus dI/dt is shown in figure 9b). This gives a value of I_{RM} = 10A.
Applying the formula for the turn-on losses:

$$W_{t(on)} = 1/2\ V_{supply}\ (I_O + I_{RM})^2 \cdot 1/(dI/dt)on$$

$W_{t(on)}$ is calculated to be: $W_{t(on)}$ = 0.4mJ

3.2 CALCULATION OF CONDUCTION LOSSES AT $T_j = 100°C$.

The value of the parameters E_o and R_o have been evaluated from the graph of I_c versus V_{ce} shown in figure 2.

$$E_o = 1.3V$$

$$R_o = 0.15 \text{ Ohms}$$

Taking I_{AVG} to be 3.5A and I_{RMS} as 5.8A, P_{on} can be calculated using the equation from section 2.2

$$P_{on} = 9.6W$$

3.3 CALCULATION OF THE TURN-OFF LOSSES.

A. High dV/dt (2500V/µs)

Using a value for $R_{g(off)}$ = 47Ohms and taking into account the dV/dt curve of figure 10 a; at a switched current of 10A the energy dissipated in turning off is

$$W_{t(off)} = 1.1 \text{ mJ/cycle}$$

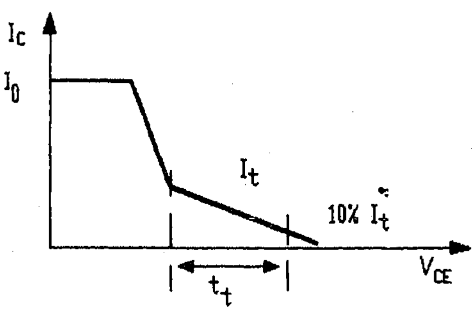


Fig. 3 - Turn-off losses

B. LOW dV/dt (500V/µs)

Using $R_{g(off)}$ = 47 Ohms and switched current of 10A again

$$Wt_{t(off)} = 0.3 \text{ mJ/cycle.}$$

Summarising these values show that the total power dissipated is dependent on the operating frequency.
Accepting that the maximum power that can be dissipated from the device at 100°C is 40 Watts for this device in a TO-220 package, it is simple to calculate that using high dV/dt the IGBT has an upper limit of operation of 20kHz while with low dV/dt operation is possible up to 40kHz.

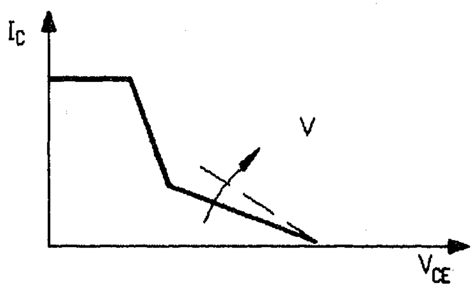


Fig. 4 - Influence of supply voltage on turn-off losses

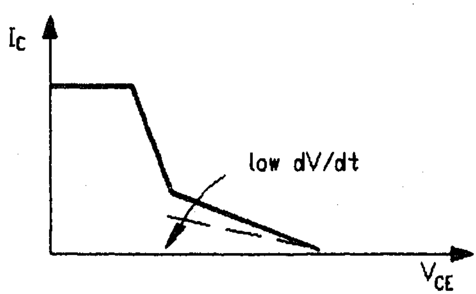


Fig. 5 - Dv/dt effect

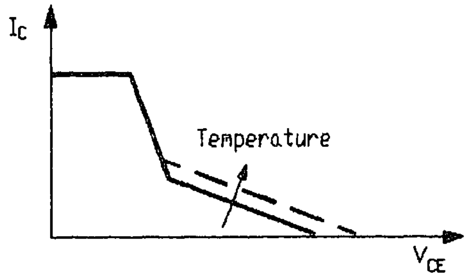


Fig. 6 - The effect of temperature on I_t and t_t

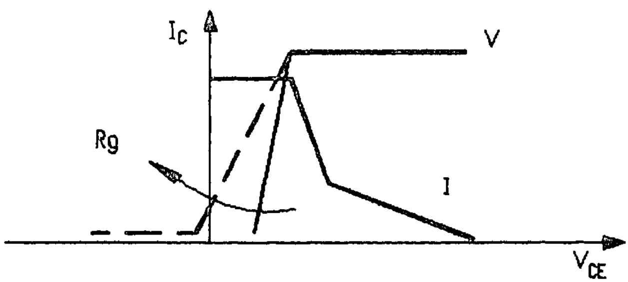


Fig. 7 - The influence of gate resistance

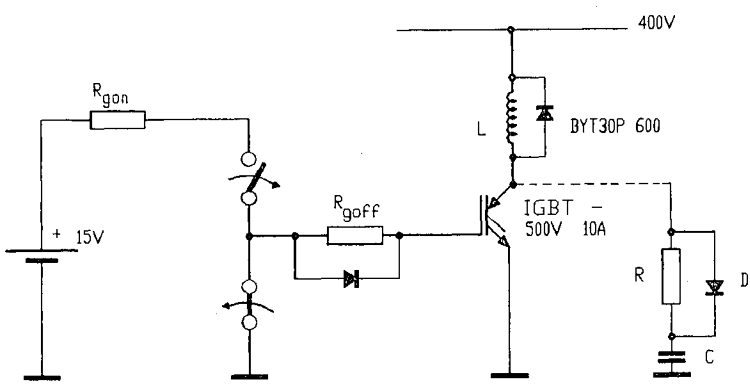
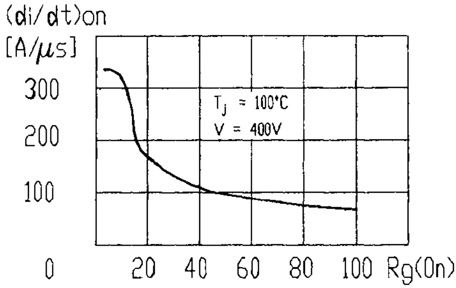
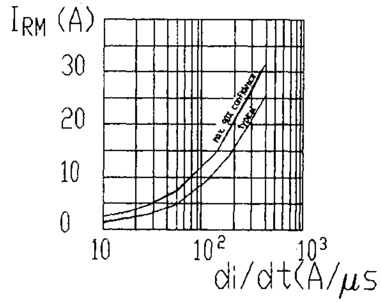


Fig. 8 - Basic circuit

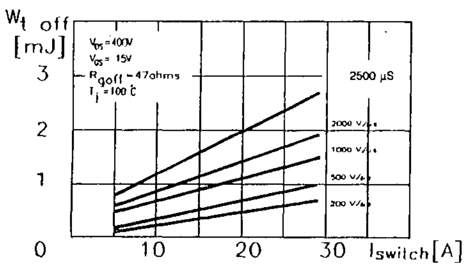


9a

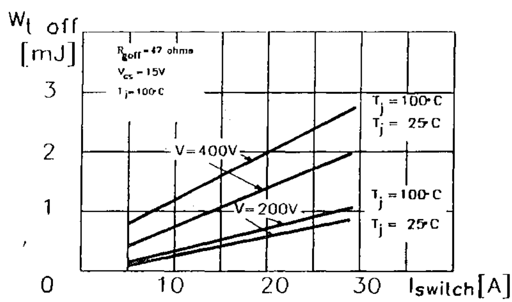


9b

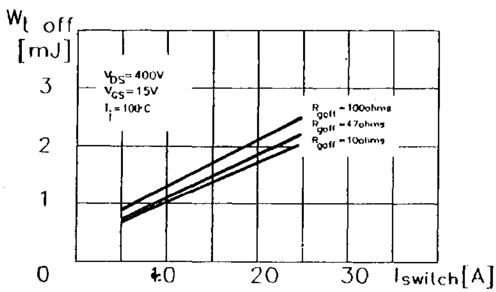
Fig. 9



(a) influence of dv/dt



(b) influence of T_j and V supply



(c) influence of Rgoff, gate resistance at turn-off

Fig. 10 (a,b,c) - Controlling factors for turn-off losses in a "2nd generation"500V/10A IGBT

	Wt(on)+Wt(off)	P(on)	Total Power Dissipated
High dV/dt	1.5mJ/cycle	9.6W	$1.5 \cdot 10^{-3} \cdot$ frequency + 9.6 W
Low dV/dt	0.7mJ/cycle	9.6W	$0.7 \cdot 10^{-3} \cdot$ frequency + 9.6 W

4. CONCLUSION

IGBTs are rugged, easy to drive and cost effective switches for high voltage chopper applications. They are capable of sustaining high current densities. Their operating frequency has been shown to be dependent on the operating conditions and a straight forward method of estimating this was discussed.

For applications such as motor drives they are robust and reliable alternatives to bipolar transistors and Power MOSFETs.