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Reduce Compact Fluorescent Cost with Motorola's PowerLux™ IGBT

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INTRODUCTION

Compact Fluorescent Lamps (CFL) are becoming more and more popular in the consumer market because they bring a significant energy savings when compared to the incandescent bulbs they are intended to replace. As a consequence, the today's focus is to cut the costs and miniaturize the electronic circuits associated with these low pressure lamps, which makes them more affordable to the consumer.

Although there are many solutions for driving fluorescent lamps, almost all the electronic ballast's for CFLs use the half bridge topology described in this paper.

CIRCUIT ANALYSIS

The CFL application can be split in two main types of products:

a – The disposable one: the attached electronic circuit is thrown away when the lamp wears out.

b – The replaceable one: a new lamp is plugged into a socket (which contains the electronic circuit), making the expected life many times longer.

The benefits and consequences of each solution is as follows:

1 – Disposable CFLs are cheaper to manufacture, but as the lamp and circuit are both replaced, turn out to be more expensive in the long term. The expected life time is 8000 hours for European models and about 10000 hours for those built for North America.

2 – Replaceable CFLs require a more rugged electronic circuit, to survive the expected longer life time and the hazards that may occur when re–lamping the module. They are more expensive to build, but can be more cost effective in the long term. The expected life time should be at least 32000 hours to make these lamps competitive when compared to the disposable alternative.

Table 1. Comparison of Topologies

	BVCES	PEAK CURRENT	COMPLEXITY	COST	DIMMING	CONCEPT
FLYBACK	1200 V	400 mA	HIGH	MEDIUM	FEASIBLE	PARALLEL RESONNANT
PUSH–PULL	1600 V	400 mA	MEDIUM	MEDIUM	DIFFICULT	PARALLEL RESONNANT, CURRENT FED
HALF BRIDGE	700 V	350 mA	LOW	LOW	EASY	SERIES RESONNANT, VOLTAGE FED

NOTES: * All Values Assume Disposable CFL Module, P_{out} = 20 W, V_{line} = 230 V.
   * Costs are Defined as Global for a Fully Assembled CFL.
   * FLYBACK = Single Power Switch.
   * PUSH–PULL and HALF BRIDGE = Dual Power Switches.

In both cases (disposable or replaceable), the ballast circuit can be designed with one of the topologies given in table #1, keeping in mind that the single switch FLYBACK is not necessarily cheaper than a two switch alternative.

The circuit we'll describe throughout this paper is based on the HALF BRIDGE topology, operating from the 230 V line.

The basic circuits given in figures #1 and #2 show the two alternatives: self oscillant or driven by a specific controller.
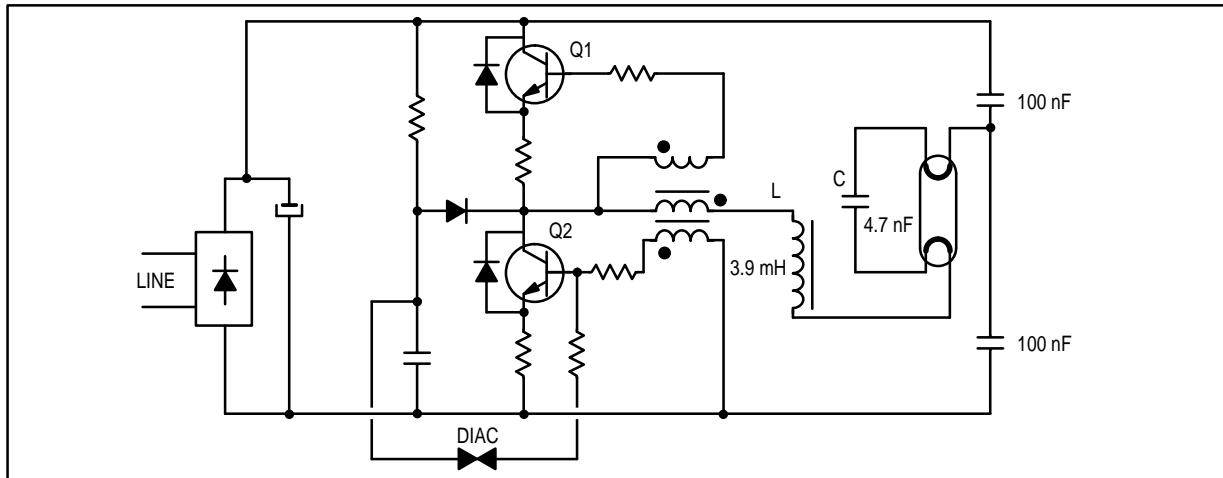
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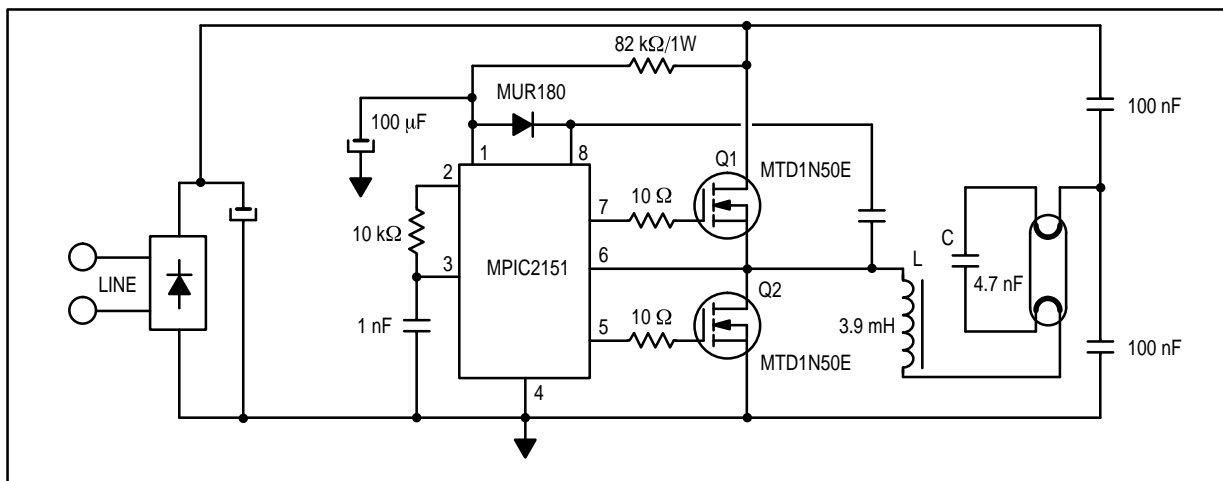**Figure 1. Basic Self Oscillant CFL**

Figure 2. Basic Driven CFL

The resonant frequency of a serial RLC network is given by equation [1]:

$$fo = \frac{1}{2 * \pi * \sqrt{(L * C)}} \qquad [\ 1\ ]$$

Out of resonance, the impedance is given by equation [2]:

$$Z = \sqrt{(R^2 + (L\omega - 1/C\omega)^2)} \qquad [\ 2\ ]$$

When the frequency equals resonance, the impedance drops to a minimum:

$$Z = R \qquad [\ 3\ ]$$

The quality factor of the RLC network is given by the equation [4]:

$$Q = (L * \omega)/R \qquad [\ 4\ ]$$

At resonance, the voltage across the capacitor is at its maximum as given by [5]:

$$V_C = V_{CC} * Q \qquad [\ 5\ ]$$

From the AC stand point, a half bridge circuit is built around two power switches and is loaded by a series RLC network. It operates in three modes:

Phase 1: Pre–heating the filaments to improve start efficiency and extend the tube's life time. The operating frequency forced by the converter moves to point A on the curve given in figure #3. The current is high enough to rapidly heat–up the filaments.

Phase 2: Striking the fluorescent lamp. The frequency is pushed close, but preferably not equal, to the RLC resonance (point B). The current increases, the voltage across capacitor C_1 is at a maximum, as given by equation 5, and the lamp strikes.

Phase 3: Continuous operation – The operating frequency is far from the RLC_1 resonance (point C) and the current flowing through the lamp is mainly dependent upon the value of inductor L (leaving aside, for the moment, the other parameters), hence upon the frequency of the chopped voltage provided by the two switches.

One must keep in mind that capacitor C_1 is being shorted by the low impedance of the fluorescent tube, the series RLC_1 circuit is heavily damped and no significant Q coefficient can be generated in the network.
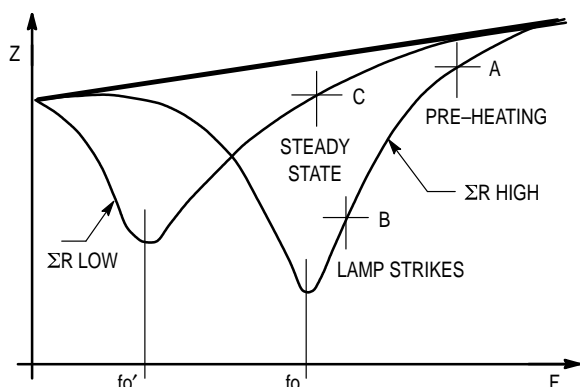
Figure 3. RLC Series Resonant Network Behavior

If we use bipolar power semiconductors to build the output stage of the converter, we must take into account the switching characteristics of these devices because they highly influence the performance of the electronic ballast. As one can imagine, a bipolar transistor based self oscillant topology is most sensitive to the AC parameters:

– The frequency **– F –** is dependent on the storage time **– t_{si} –** as depicted in figure #4.

– The current fall time **– t_{fi} –** and the storage time **– t_{si} –** are highly sensitive to the reverse bias applied across the Base/Emitter.
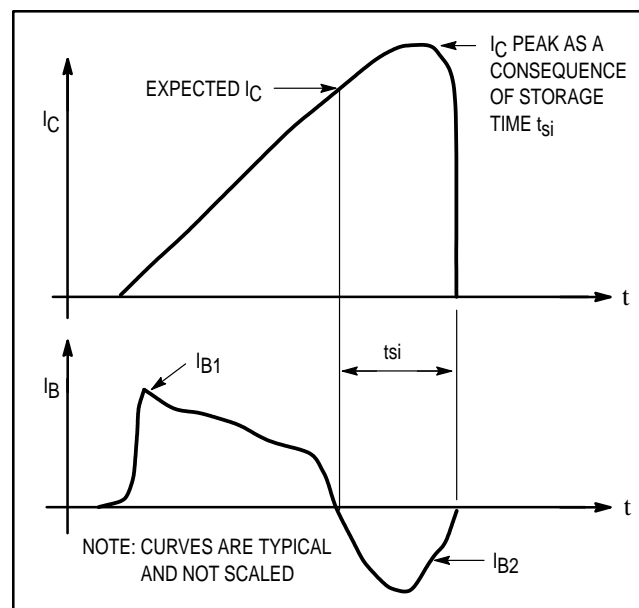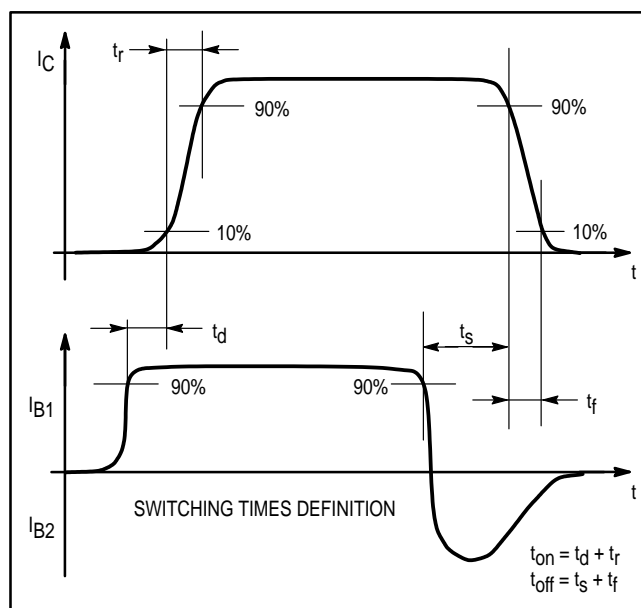
For a standard bipolar transistor, the storage time is dependent on the gain (β) of the transistor associated with the operating point I_C and the drive conditions I_{B1} and I_{B2}. The H2BIP device is free from this relationship, for a given I_C range, and provides a far better design than older bipolar technology.

Typical I_C and I_B current waveform in a self oscillant half bridge circuit.

Figure 4. Switching Characteristics of a Bipolar Power Transistor

With regard to a MOSFET, storage time can be disregarded since, like the fall time, it's mainly dependent upon the impedance connected across Gate/Source at turn off. At first glance, the MOSFET would be preferred for these types of electronic ballast, but it presents two major drawbacks:

– The Rdson is a function of the Breakdown voltage capability: the higher the BVDSS, the higher the Rdson for a given chip size.

– The Rdson has a positive temperature coefficient and its absolute value will double over the usual +25°C to +125°C temperature range, yielding high on state losses into the silicon.

Table #2 gives a summary of the comparison between BIPOLAR and MOSFET power devices. One must point out that, for low line voltage (eg. 120 V), the MOSFETs can be the preferred choice because they bring more advantages than drawbacks compared to other semiconductors technologies.

Table 2. Bipolar to MOSFET Simplified Comparison for 500 V Breakdown Voltage

	V_{ON}	GAIN	tsi	tfi	RELATIVE CHIP SIZE
BIPOLAR	$V_{CEsat} < 1$ V	$\beta = f(I_C)$ $\beta = f(T_J)$	1.8 µs <—> 2 µs* 1.5 µs <—> 5 µs**	120 ns <—> 400 ns	1.0
MOSFET	$R_{DSon}*ID = f(T_J)$	$g_{fs} =$ CONSTANT	200 ns^x	40 ns <—> 80 ns	1.2

NOTES: * Bipolar H2BIP Technology with Built–in Free Wheeling Diode
    ** Standard Bipolar Devices
    x Dependent Upon the G/S Impedance

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On the other hand, since the Rds$_{on}$, hence the ON losses, is a function of the die size, the designer must cope with the cost/performance ratio to make the final design cost effective.

To overcome the compromise associated with either the bipolar transistor or the MOSFET, we have developed a new concept, at the silicon design level, yielding an IGBT fast enough to be suitable for the 40 kHz operating frequency commonly used in the CFL electronic ballast's.

PowerLux™ IGBT

Main characteristics

The IGBT technology is widely used for high current/high voltage applications, such as motor controls and automobile engine ignition circuits. It has proven to be rugged and reliable, but is prone to high switching losses when operated above a few kilohertz, due to the large tail current that appears at turn off.

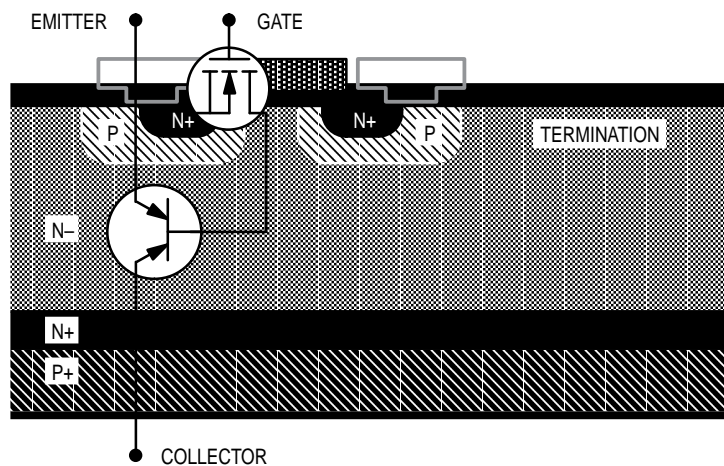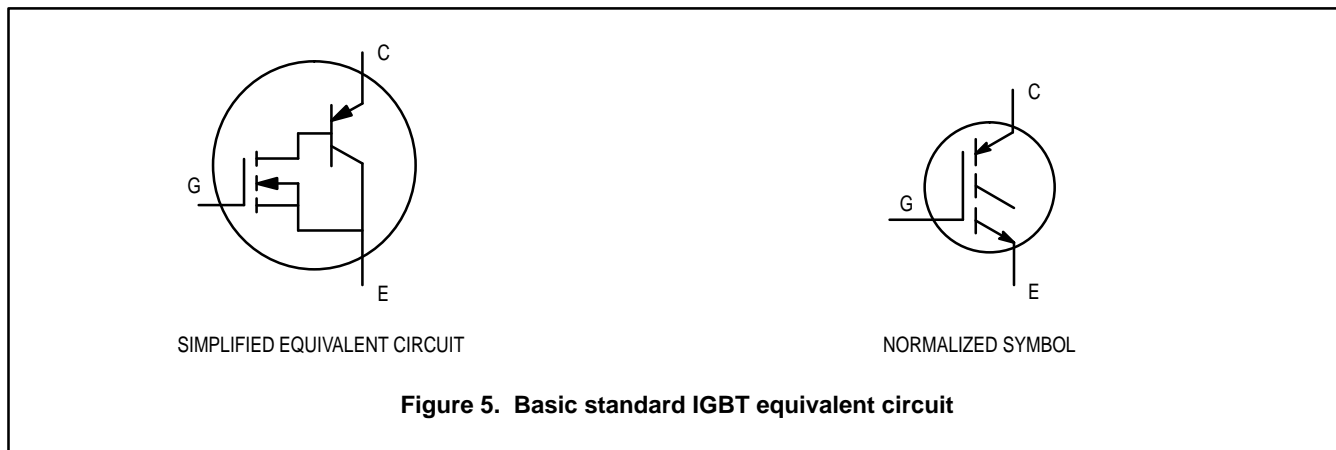Figure #5 gives the basic equivalent schematic of a standard IGBT, the die cross section is shown in figure #6.

SIMPLIFIED EQUIVALENT CIRCUIT

NORMALIZED SYMBOL

Figure 5. Basic standard IGBT equivalent circuit

Figure 6. Typical Standard IGBT Die Cross Section

The basic equivalent circuit shows that the IGBT is a MOS gated device which controls the base current flowing into a junction transistor which does not have an integrated free wheeling diode.

This concept has two advantages:

1 – The input is identical to a MOSFET: high impedance, voltage driven.

2 – The output is equivalent to a BIPOLAR: Von is identical to Vcesat and, unlike the MOSFET, does not increase, for a given current I$_D$, over the operating temperature range.

On the other hand, since there is no path to remove the extra charge stored in the Base/Emitter of the output stage when the

device switches off, the drain current will exhibit a long fall time known as a tail current effect. As a matter of fact, during switch off, the MOS turns off much faster than the bipolar junction and the charges are trapped in the Base structure. This is the main drawback that limits today's standard IGBT to applications running in the 20 kHz range.

However, the technology is robust and, thanks to its MOS input, far easier to control than a Bipolar transistor. These advantages make the IGBT the best choice to build high current drivers as stated above.

The PowerLux™ IGBT overcomes this speed limitation and provides a prime solution to design a CFL electronic ballast as described in chapter 3.

IGBT Main Electrical Characterisitics

Like any other semiconductor, the IGBT's electrical Characterisitics are limited by normalized DC and AC parameters that are defined in the designer's data sheet. Although all these parameters are important, the critical ones, from an electronic ballast stand point, are linked to the expected behavior of the device when it's operated in the application:

1 – Operating frequency stability

2 – Turn – OFF losses

3 – ON state losses

4 – Voltage capability

The first two fields are dependent upon the switching performance, the third one being a function of the transconductance and associated ON voltage, the fourth being dictated by the breakdown voltage of the device.

The IGBT has two main drawbacks:

1 – Latching effect: the device latches when the I_D current exceeds a given value.
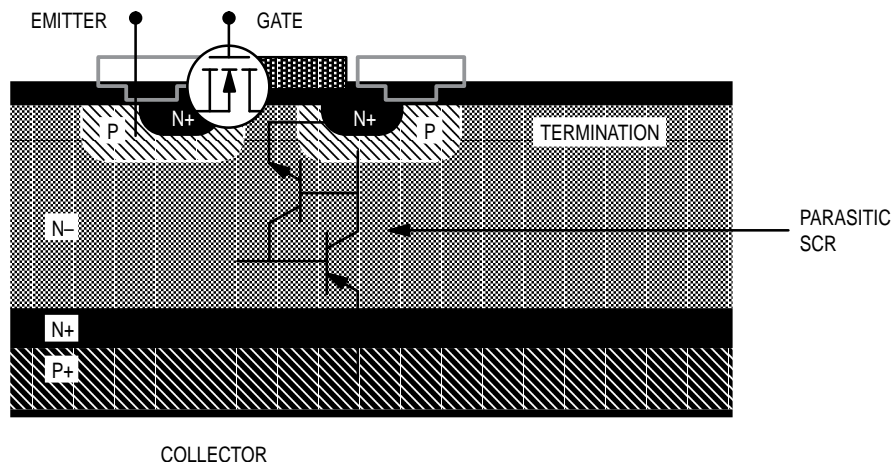
2 – Current tail.

Figure 7. Parasitic Thyristor into the IGBT Structure

The tail current is of prime importance for the electronic ballast because it generates most of the switching losses. As a matter of fact, since these applications operate, in steady state, in the 30 kHz to 50 kHz range, this type of circuit cannot accommodate semiconductors with a poor switching performance: very fast current fall times are mandatory as described figure #8.

The tail current and therefore, the switching loss is the most important parameter because it determines the suitability of the IGBT for operating at 40 kHz.
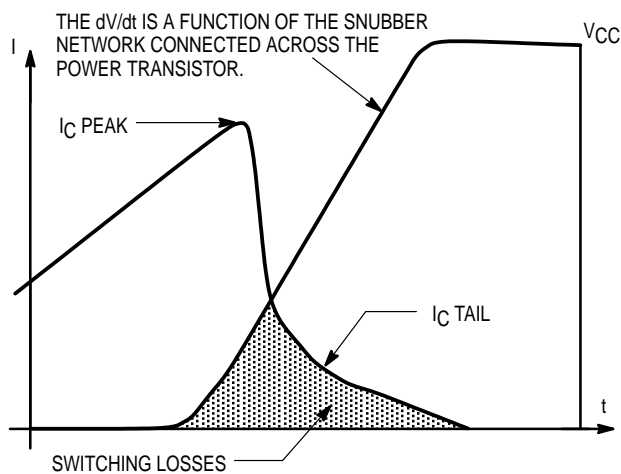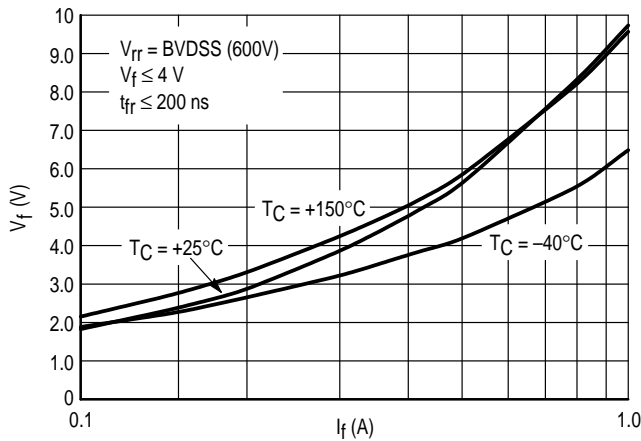
As shown by the typical circuit diagrams given in figures #1 and #2, one must provide a path to re–circulate the inductive current at the turn off of either switch. When the output stage is built with Bipolar devices, an extra free wheeling diode is connected across the Collector Emitter. The H2BIP* technology was developed three years ago to integrate this function into the power die. Of course, one can avoid the diode, using the Base Collector junction to accomplish the function, but this is not recommended for two reasons:

1 – If the voltage dropped across the Base/Emitter external network is larger than the BV_{EBO}, then the junction is avalanched on each pulse and the expected life time can be downgraded if the energy is higher than the maximum rating of the transistor.

2 – Forcing a forward current into the Base/Collector junction yields poor control of the switching performance (unless the transistor is operated in a very hard saturation mode, but this may lead to high losses in the drive network), which, in turn, downgrades the RBSOA characteristics of the device (a characteristic that allows the transistor to sustain the fault operating conditions).

The MOSFET alternative brings the advantage of the body diode that performs the free wheeling function.

Until now, the IGBT did not have an intrinsic diode. The PowerLux™ devices have being designed with a monolithic diode, making the device suitable for a CFL electronic ballast applications. The main characteristics of that diode are summarized in table #3.

Figure 8. Switching Losses

Table 3. PowerLux™ IGBT Diode Characteristics

V_{rr} = BVDSS (600V)
$V_f \leq 4$ V
$t_{fr} \leq 200$ ns

T_C = +150°C
T_C = +25°C
T_C = −40°C

PowerLux™ IGBT Electrical Performance

The fast IGBT, specifically designed for the compact Fluorescent Lamp applications, brings an integrated collector/emitter diode and has switching capabilities that make the device suitable for operating frequencies up to 40–50 kHz. This new part belongs to the PowerLux™ family, the basic equivalent circuit being given in figure #9.

SIMPLIFIED EQUIVALENT CIRCUIT

NORMALIZED SYMBOL
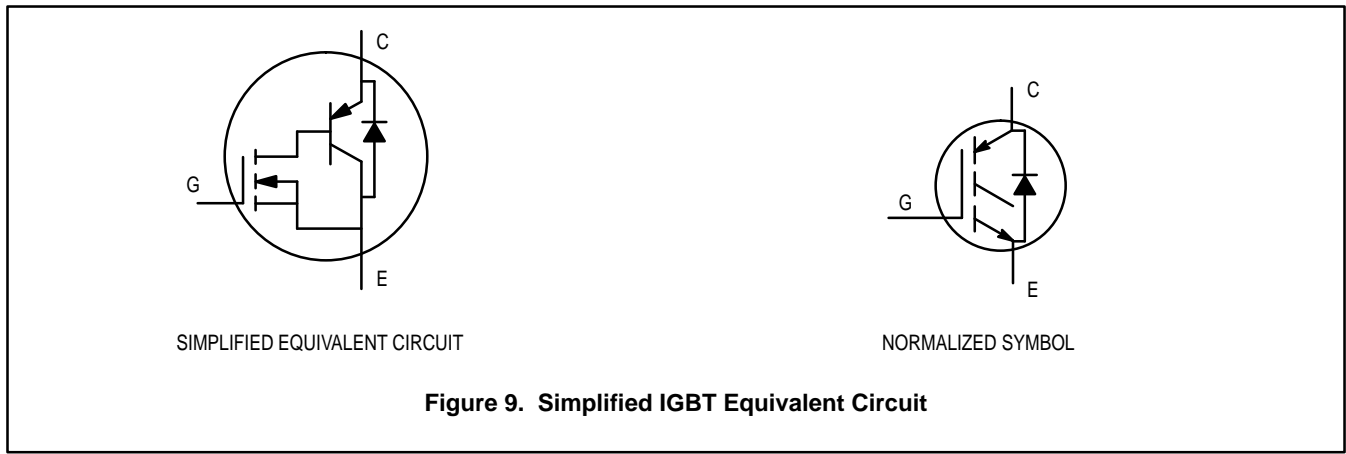
Figure 9. Simplified IGBT Equivalent Circuit

BREAKDOWN VOLTAGE BV$_{DSS}$

Since the device is intended to be used on the 230 V mains, with no or limited voltage regulation, the BV$_{DSS}$ has been targeted at 600 V minimum. Since that parameter has a positive temperature coefficient, as shown by the curve given figure #10, the BV$_{DSS}$ is free of thermal run away.
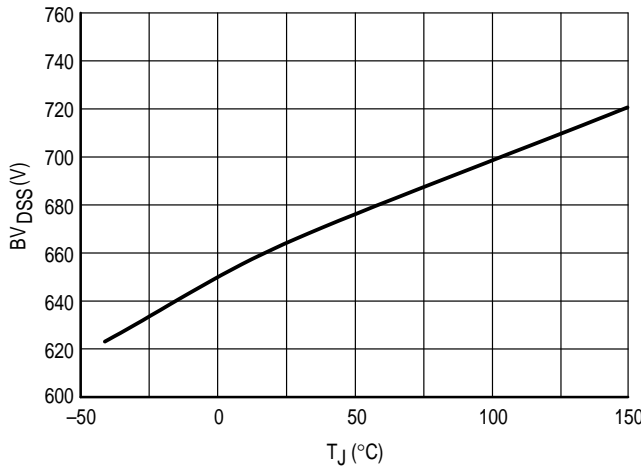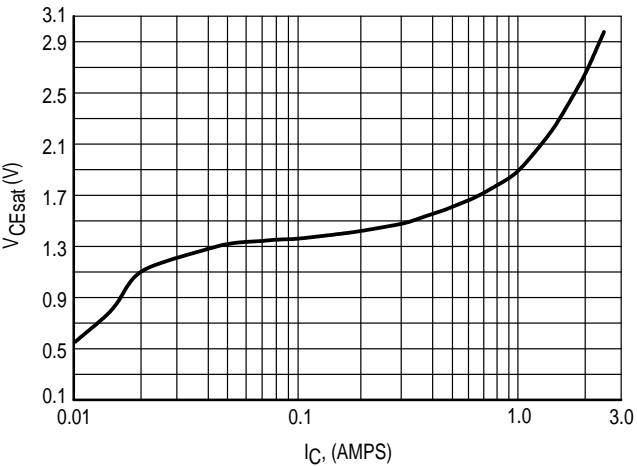

Figure 10. BVDSS = f(Tj)

Figure 11. V_{ON} = f(I_D) @ Vgs = 10 V, T_C = +25°C

ON STATE VOLTAGE V$_{ON}$

This parameter plays a significant role since, associated with the collector current I_C, it generates the ON state losses which, together with the switching losses, determine the junction temperature for a given junction to ambient thermal resistance and ambient temperature.

Keeping in mind the CFL as the target application, the IGBT has been designed to get a $V_{ON} \leq 1.80$ V at $I_C = 400$ mA. The curve given figure #11 shows the typical value V_{ON} as a function of the collector current.

As depicted in figure #12, the V_{ON} has a negative thermal coefficient and there is no risk of thermal run away when the device is operated within its maximum ratings.
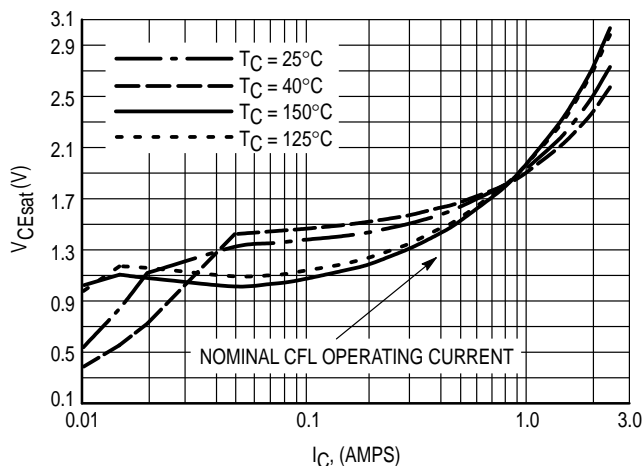

Figure 12. $V_{ON} = f(T_J)$, Vgs = 15 V

THRESHOLD VOLTAGE Vgs_{th}.

The threshold voltage is defined as the Gate/Source voltage that yields a 1 mA collector current. For an electronic ballast application, one must consider two main types of circuits:
1 – Self oscillant: the drive voltage, derived from either a second transformer or from the output inductor, is almost sinusoidal and the Vgs_{th} spread may influence both the turn–on and turn–off, hence the operating frequency. Like the MOSFET, this behavior is highly dependent upon the circuit used to drive the devices.
2 – IC Driven: The Vgs is supplied by a specific driver, like the MPIC2151, with a square waveform. The Vgs_{th} variation has no influence on the timing as long as it's well below the 10 V to 18 V supplied by the integrated driver. On the other hand, the minimum Vgs_{th} must be higher than the maximum logic zero level voltage sourced by the driver, otherwise the IGBT could be operated in continuous conduction mode.

Additionally, the IGBT must be made insensitive to the electrical noise coming from the very fast dI/dt and dV/dt generated in the circuit. Taking into account these specifications, the PowerLux™ IGBT has been designed with a typical Vgs_{th} of 4.0 V, the spread being ±1.00 V.

GATE SOURCE BREAKDOWN VOLTAGE BV_{GSS}

The device is protected by two back to back zener diodes, integrated into the chip and connected across Gate–Source. This clamps the Gate voltage to typically ±21 V, making the device ESD resistant (Human Body Model). However, it's recommended to use the standard procedures when handling these MOS gated devices to avoid the damage associated with high ESD voltages.

SWITCHING LOSSES E_{off}

Since one cannot fully define the behavior of the IGBT with a collector current fall time **– tfi –** only, the devices are characterized by a switch off energy **– E_{off} –** parameter which takes into account the complete tail current. The PowerLux™ IGBT has an $E_{off} \leq 5\mu J$ at Tj = +25°C, with a positive temperature coefficient as depicted in figure #13.
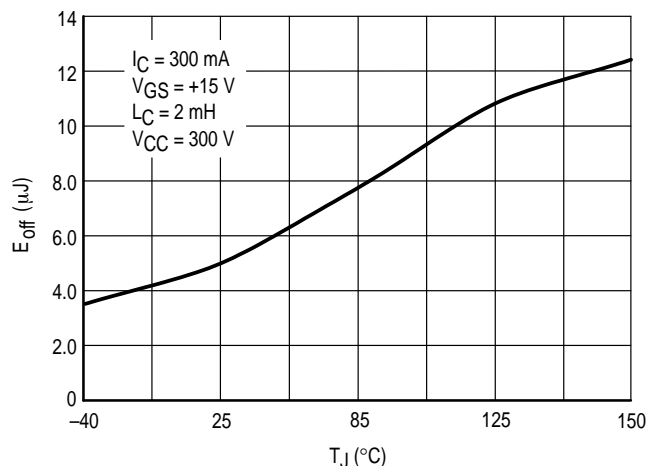

Figure 13. Switching Losses –E_{off}– as a Function of Temperature

FREE WHEELING DIODE

The built–in diode has been designed to match the CFL's requirements:

Vf ≤ 4.00 V	@ If = 400 mA, Tj = +25°C
tfr ≤ 200 ns	@ If = 400 mA, Tj = +25°C

Since the current flows into the diode for a limited duty cycle (typically 15%), the relatively high Vf does not generate more than a few milliwatts of loss into the silicon as given by equation [7]:

$$P_{diode} = If_{rms}*Vf*DC \qquad [7]$$
$$P_{diode} = 0.090*4*0.15$$
$$P_{diode} = 54 \text{ mW (maximum)}$$

PowerLux™ IGBT APPLICATION

Typical CFL circuit

With a current capability of 500 mA in steady state, up to 2 A in pulse mode, the MMG05N60E is well suited for the full power range of CFL applications. As a matter of fact, since the power into a fluorescent tube is a function of its length, hence the Von across the two electrode ends, the current flowing into the transistors is almost constant over the 7 W to 23 W range. The main difference is the strike voltage (the longer the fluorescent tube, the higher this parameter) which, in turn, needs a higher current during the start–up. Although the breakdown capability of the power device is not related to the strike voltage required by the lamp, the silicon must be able to sustain the inrush current during both the filament pre–heating and the strike sequence. The demoboard designed in the Application Lab takes care of these constraints, using the integrated MPIC2151P driver to control a couple of IGBT in the popular half bridge topology as described figure #14.
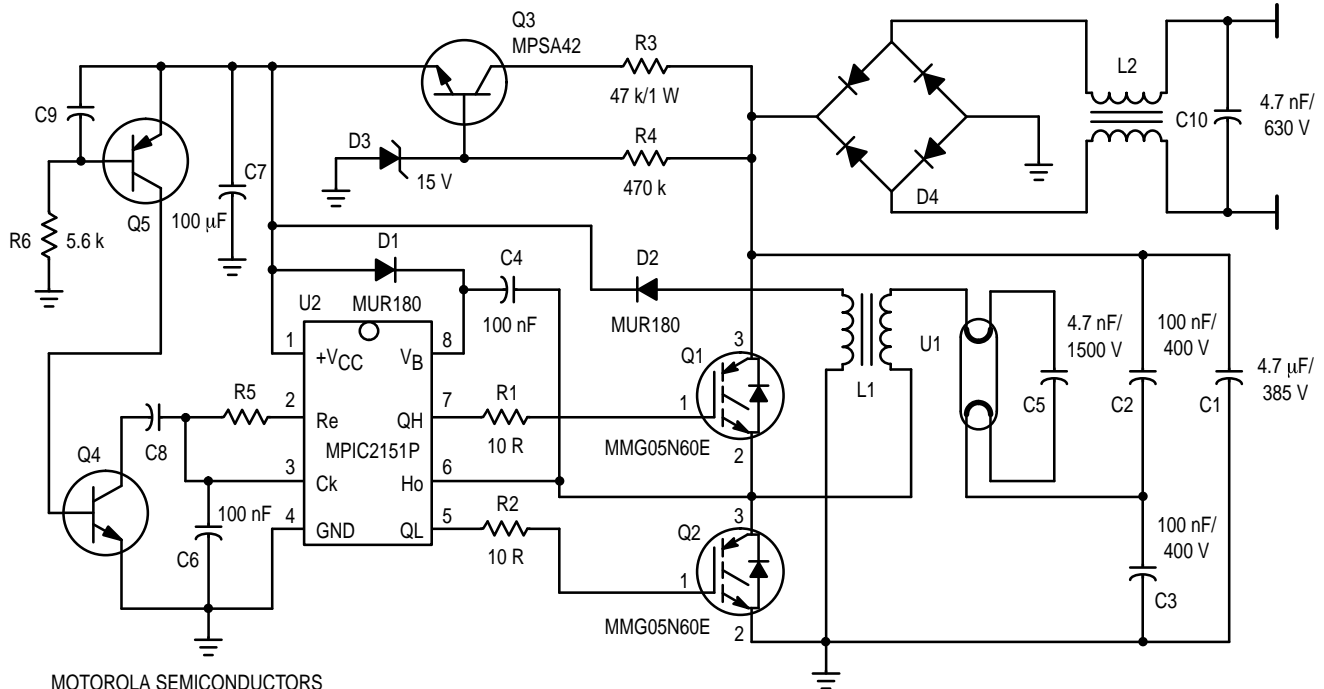
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Figure 14. Demoboard Schematic Diagram

Start–Up

In order to increase the life time of the lamp, the filaments are pre–heated prior to applying the strike voltage across the electrodes. This is achieved by starting the circuit at a high frequency, connecting capacitor C2 across the timing network after a delay coming from R5/C7. The curves given figure #15 show the behavior of the RLC series resonant network built with the components show figure #14, assuming the cold resistance of the filaments being 10 Ω each, with a thermal time constant of 500 ms.

Since this CFL is of the disposable type, there is no need for safety circuits as one should include in the linear tube used in the industrial application.
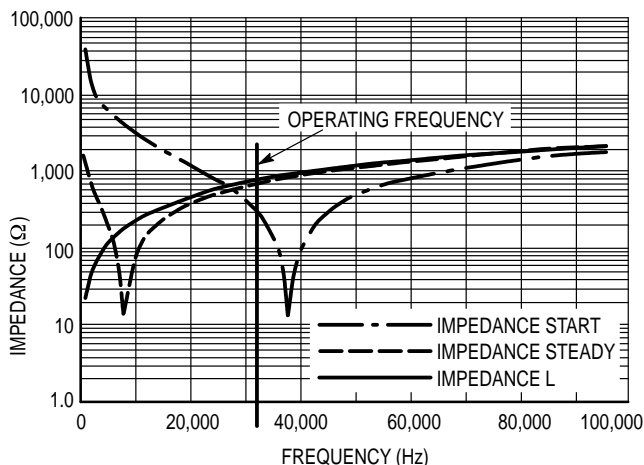
**Figure 15. RLC Series Resonnant
Circuit Typical Behavior**

Impedance START = impedance of the start–up network built with the output inductor L and resonant capacitor C.
Impedance steady = steady state impedance of the output inductor associated with the passive side of the half bridge.
Impedance L = impedance of the pure output inductor.

Steady State

Once the lamp is ON, the current is limited by inductor L, according to equation [8]:

$$I = (Vcc–Von)/L\omega \qquad [8]$$

with: Vcc = DC bus supply voltage

Von = ON state voltage across the fluorescent tube

Since both Vcc and Von are constant (for a given power and line voltage), and the operating frequency is bounded by low and high limits, the range of potential inductance values is very restricted, and L can be calculated straightforward by combining equation [9] with re–arranged equation [8]:

$$P = Von*Irms \qquad [9]$$

$$L = [Von*(Vcc–Von)]/(P*2*\pi*F) \qquad [10]$$

For an 11 W lamp powered from a 230 V nominal line, operated at 35 kHz, the inductor is:

$$L = [50*(310–50)]/(11*2*\pi*35^{+3})$$

$$L = 3.09 \text{ mH}$$

As we can observe on the curve given in figure #15, the circuit operates far from the resonance generated by L and C5/C6 in parallel. It's the designer's responsibility to select the operating point either close or far from a second resonant pole, keeping in mind the following points:

a – When operated far from resonance, the circuit is less sensitive to the tolerances of either component, but the switching losses are maximum since the power devices are switched off at peak collector current.

b – When running in a quasi resonant mode, the current is sinusoidal and switched off during the negative going slope of the sine wave. Consequently, the switching losses are minimum, but the circuit becomes more sensitive to the electrical tolerances of the passive and active components.
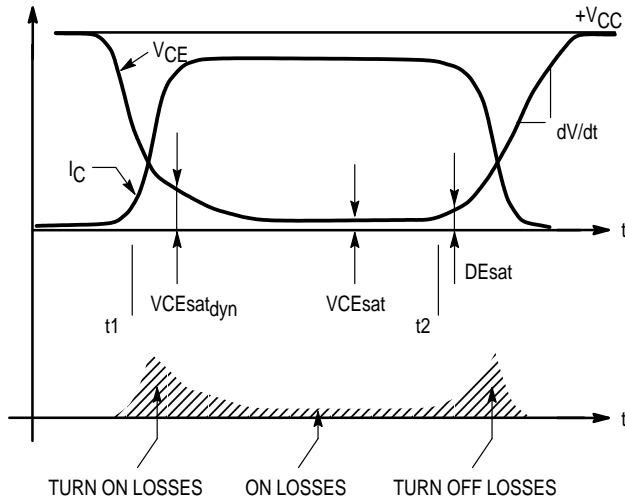
Figure 16. Switching Time and Loss Definitions

The switching losses generated by the IGBT are calculated with equation [11] and added to the ON state losses, with those coming from the integrated free wheeling diode, to derive the maximum junction to ambient thermal resistance as stated by equation [18] (see figure #16 for the definitions of the related parameters).

$$S_{WL} = E_{off} * F \qquad [11]$$

with: F = operating frequency

Eoff: energy dissipated during switch off

$$E_{off} = Vce*Ic \qquad [12]$$

The E_{off} parameter is provided by the IGBT designer's data sheet. It is fully characterized for preferred operating conditions and a set of curves gives the typical behavior as a function of the collector current and the junction temperature.

$$P_{ON} = \int_{t1}^{t2} V_{ON} * I_C * dt \qquad [13]$$

Since there is no dynamic Vcesat associated with an inductive load, equation [13] can be simplified as:

$$P_{on} = 1/2*Von*Ic_{peak}*DC \qquad [14]$$

On the other hand, one cannot abruptly cancel the current into the inductor so the integrated free wheeling diode provides a path to re–circulate this current. The forward drop of the diode is the most important parameter because it generates most of the diode losses as given by equation [14]:

$$P_D = Vf*If*F \qquad [15]$$

Like the Pon losses, equation [15] can be simplified to:

$$P_D = 1/2*Vf*Ifpeak*DC \qquad [16]$$

Thermal Analysis

The junction temperature of a semiconductor, operated under steady state conditions, is given by equation [16]:

$$Tj = Tamb + \Sigma P*Rth_{j-a} \qquad [17]$$

The maximum junction temperature for a silicon device, assembled into a plastic package, is +175°C, but in order to improve the long term reliability, it's highly recommended not to continuously operate the device above +150°C. Also, if the silicon temperature is forced to that value, the pins of small packages like the SOT223 or the TO92, will be nearly at the same temperature as the junction and the solder attach to the printed circuit board will be rapidly downgraded. Consequently, it's far better to operate the power devices around +120°C junction temperature to avoid the risk of long term overall reliability. Assuming the maximum ambient temperature being +90°C, we can calculate the maximum junction thermal resistance by re–arranging equation [16] as stated below:

$$Rth_{j-a} = (Tjmax - Tamb)/\Sigma \qquad [17]$$
$$Rth_{j-a} = (120-90)/0.30$$
$$Rth_{j-a} = 100°C/W$$

Since the SOT223 junction to ambient thermal resistance is 156°C/W, it's obvious that it cannot be operated without a minimum heatsink to reduce the Rthj–a to 100°/W. This is easily achieved by using extra copper area on the pcb to mount the IGBT. The curve given figure #17 gives the thermal resistance as a function of the foot print area. A single copper rectangle (or square) of 10x6 mm (using standard 35 μm thickness) is enough to get the expected Rth_{j-a} as calculated above.
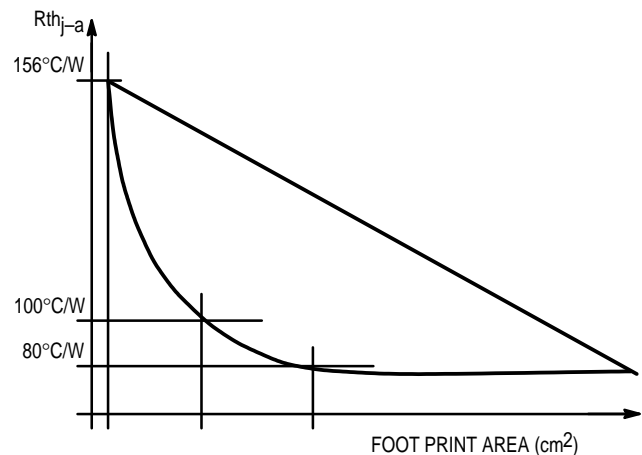

Figure 17. Junction to Ambient Thermal Resistance of a SOT223 Package

Note: the curve is given for information only: accurate data are available from the Motorola surface mount device technical booklet.

Engineering Test Results

The engineering tests, performed in our Toulouse Applications Laboratory, are summarized by the oscillograms given figures #18 and #19. The critical point is, beside the V_{on}, the collector current fall time: as depicted in figure #19 the current tail is negligible and the PowerLux IGBT can safely operate up to 50 kHz in the 5 W to 23 W power range.
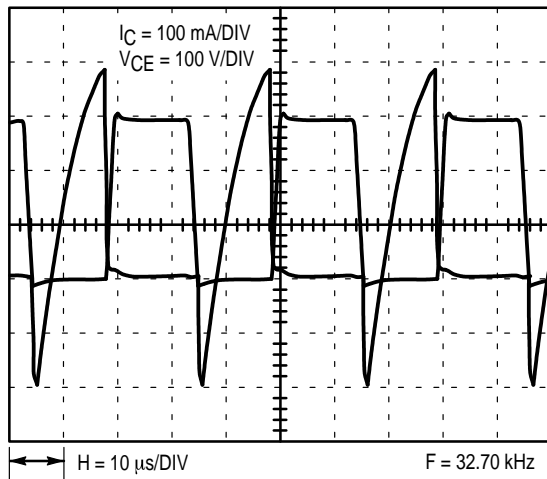
Figure 18. Steady State Operation (V$_{line}$ = 230 V)

Figure 19. Switching Losses in the Steady State

In the steady state operation, the case temperature of the TO92–1W device increases by +26°C, yielding the ability to operate the IGBT up to +100°C ambient. Since, unlike the bipolar device, there are no risks associated with the degradation of the hfe with the temperature, the PowerLux™ IGBT is free from thermal runaway as a result of desaturation. On the other hand, thanks to a latch up trip current level above 5 A, the IGBT is safe from such mechanism, even under start–up conditions.

Low Cost Version

The schematic given figure #14 can be simplified to reduce overall cost of the CFL as depicted figure #20:

a – Replace the pre–heating active network with a PTC and an extra high voltage capacitor in the resonant pole. The

PTC can also be connected directly across C5 to dump the circuit until the PTC reaches its high temperature value.

The drawback is a loss of control and the heat generated by the PTC. As a matter of fact, to get the expected resistance, the PTC must operate at +130°C under steady state conditions.

b – Remove the active VDD supply, using a power resistor R3. Like the PTC, the drawback is mainly the heat coming from a component which dissipates around 1 W.

c – Use one capacitor only to close the loop on the passive side of the half bridge. At that point, one must pay extra attention to the imbalance of the magnetic circuit together with the RFI generated by the module.
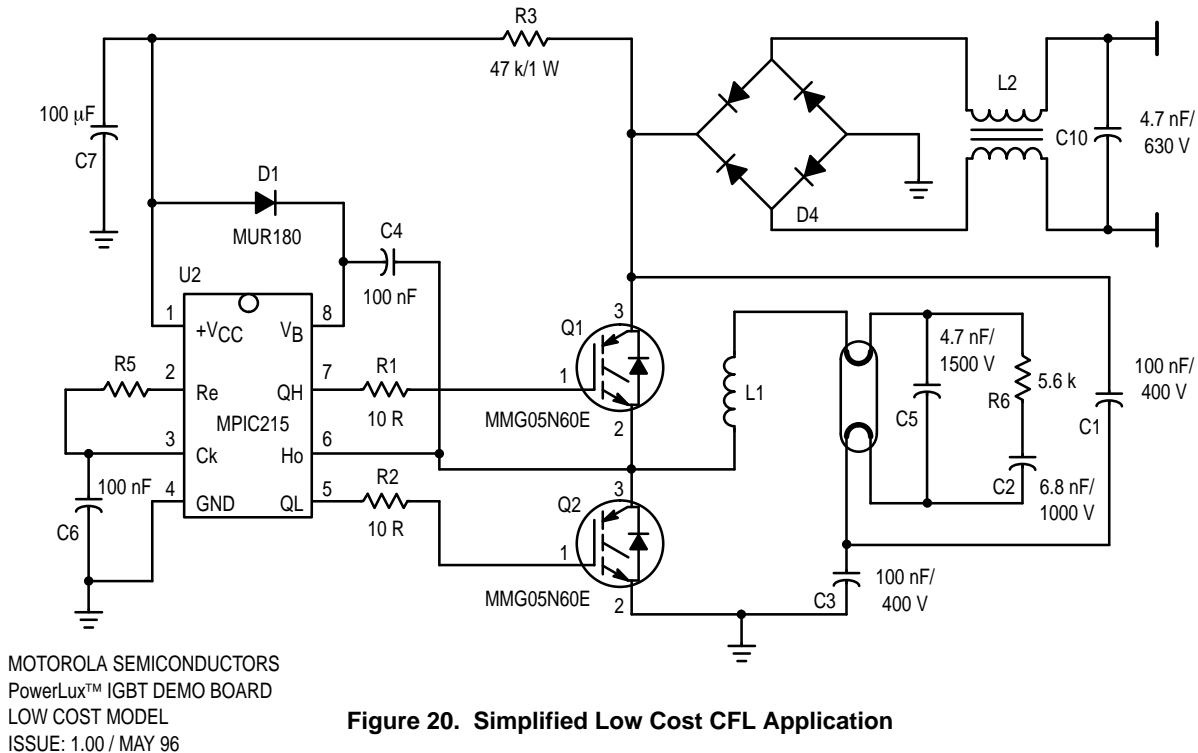

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PowerLux™ IGBT DEMO BOARD
LOW COST MODEL
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**Figure 20. Simplified Low Cost CFL Application**

The cost can be further reduced by using a self oscillant circuit as depicted figure #21. Transformer T1 provides the positive feedback to drive the IGBT, the associated components

R,D,C being used to improve the dynamic performance of the power device. The drive can also be derived from the output inductor, but such analysis is beyond the scope of this paper.

Figure 21. Self Oscillant, Low Cost CFL Basic Circuit

One must point out that, although this is the lowest cost solution, it is not easy to design, and it has been already proven that a driven circuit (with a dedicated driver), is usually faster to design and, consequently, the best solution for a short time to market. Also, a driven circuit can be easily modified to control lamps of different output power.

Typical PFC Circuit

Nowadays, European Regulations specify that any piece of electronic equipment connected to the line must have a cosΦ of 0.94 minimum with a THD of 25% maximum for the third harmonic. However, the full implementation of that regulation for input power below or equal to 30 W, hence for the CFL, has been postponed to 1998 and, at the time of printing, no decision has been taken.

In any case, since the PFC will eventually be mandatory, a typical application circuit is given in figure #22.

Figure 22. Basic Low Cost Power Factor Correction Circuit
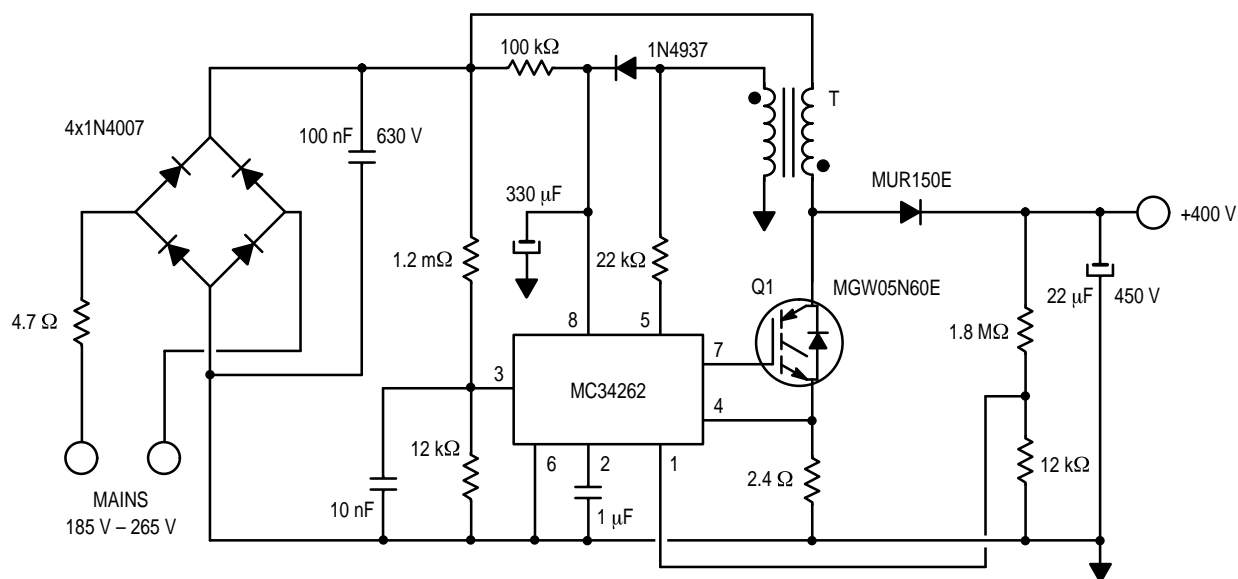
Since the PFC boost circuit can easily operate in the 40 kHz range, with a peak current below 500 mA, the MGW05N60E PowerLux™ IGBT can be used to fulfill that particular function, bringing another net cost saving to the CFL and other low power, off–line applications.

Device Comparison: BIPOLAR – MOSFET – IGBT

Selecting a power semiconductor to design an electronic ballast is not straightforward because each of the available technologies has advantages and drawbacks. The choice is the result of the end product specifications and the compromises the design engineer can make for a given application.

Table #4 below is intended to aid the designer in selecting the power device for either a Compact Fluorescent Lamp or a Linear Industrial Tube ballast.

Table 4. Power Semiconductor Technologies Comparison

DEVICES	Vsat	BVCEO	FBSOA	RBSOA	Storage time tsi	Fall time tfi
BIPOLAR standard	LOW	HIGH (1600 V max)	2nd BV limited	Vbe bias dependent	High, wide spread	Medium
BIPOLAR H2BIP	LOW	HIGH (1800 V max)	2nd BV limited	Vbe bias dependent	Low, narrow spread	Fast
MOSFET	=Rdson*ID	Medium (600 V max)	Square =BVDSS	Square =BVDSS	negligible	very fast
IGBT PowerLux™	Medium	Medium (600 V max)	2nd BV limited	Gate bias dependent	negligible	fast

DEVICES	Die size (relative)	Smallest package	Operating Tj °C max	DRIVE	Relative Cost	Dimming sub function
BIPOLAR standard	1.00	DPAK	+175°C	IB1 – IB2 Complex	Medium	Complex
BIPOLAR H2BIP	1.15	DPAK	+175°C	IB1 – IB2 easy	Medium	Medium
MOSFET	1.30	DPAK	+175°C	Voltage Vgs very easy	High	Easy
IGBT PowerLux	0.40	TO92–1W SOT223	+175°C	Voltage Vgs very easy	Low	Easy

CONCLUSION

The PowerLux™ IGBT meets all of the electrical requirements for a Compact Fluorescent Lamp application. Thanks to its chip size which can be packaged in either a SOT223 or a more conventional TO92. This brings a significant cost savings, compared to other semiconductor technologies, without downgrading the global performance of the circuit.

Because it is a MOS gated device, it can be driven by standard MOS drivers, allowing the design of high end modules with built–in functions like dimming or remote control, functions not easily achievable with a bipolar transistor.

The next step is the development of a more powerful device to fit the 55 W and 120 W applications.

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