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Motorola's Solutions for Very Low Power in Standby Mode in Switchmode Power Supplies

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The following application note summarizes the Motorola solution to achieve a very low power in standby mode

MC33363B keeps standby-power below 50mW on universal mains

The quest for low standby power represents one of the most challenging task for Switch-Mode Power Supply (SMPS) designers. This sentence becomes pregnant when you tackle very low output loads: the efficiency degrades to poor values and you barely reach 45% for outputs as low as 300mW. Thanks to the appropriate technology, the below circuit helps you easily reach 34.8mW when the load is disconnected ($V_{in} = 85VAC$). The consumption increases up to 44mW at high line (260VAC). The efficiency also exhibits a good value at low loads: 67.45% @ $P_{out} = 880mW$ and $V_{in} = 120VAC$.

A clever way to build such a low standby-power SMPS is to implement an hysteretic architecture. The explanation is very simple: if the output voltage stays below the target, the controller issues high-frequency pulses until the target is reached. When this condition occurs, the controller brakes the pulse train and the switch is kept fully OFF until the output voltage drops below the threshold. Then a new burst of pulses takes place and so on. In the lack of load, the controller thus delivers packet of pulses that simply compensate the self-discharge of the output capacitor. Just very few energy is then necessary to keep the output alive and the repetition rate of the pulse packet is long. You obviously draw insignificant power from the source.

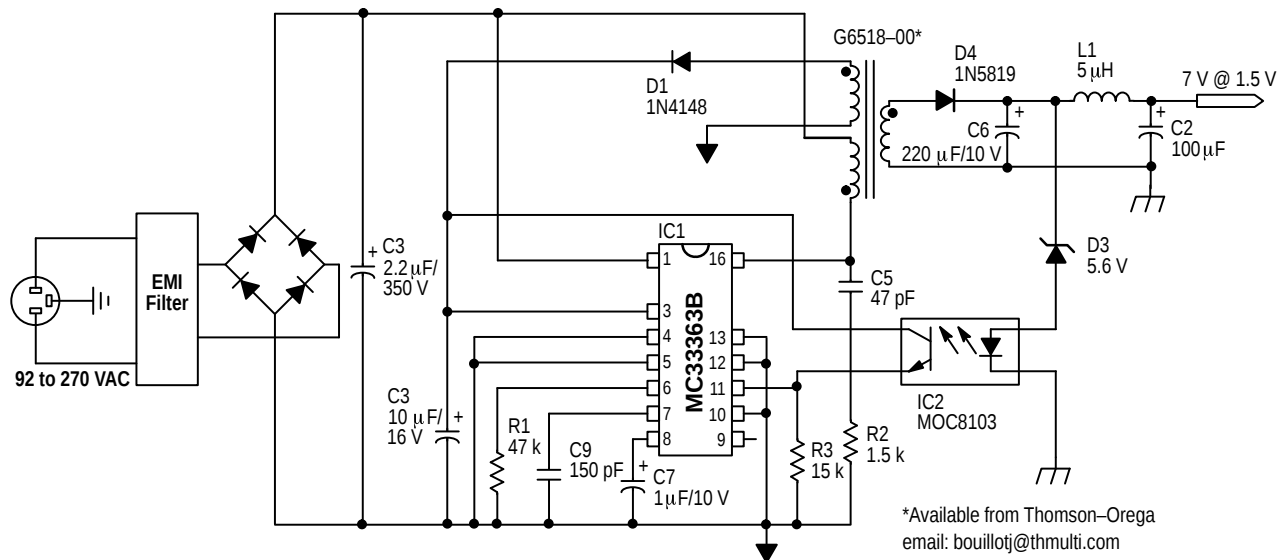


Figure 1. A MC33363B associated with a low-cost optocoupler helps you reach ultra-low standby power.

Figure 1 depicts the unusual implementation of the circuit using an MC33363B from MOTOROLA (Phoenix, AZ). This circuit includes a 700V/1.0A senseFET PowerSwitch and a startup current source. The novelty consists in using the quick OverVoltage Protection (OVP) pin combined with the feedback path. When the OVP level is reached, the internal MOSFET is stopped, leaving the output slowly discharging. When the OVP signal drops below its threshold, the MOSFET is turned on again: the circuit regulates in an hysteretic mode, exactly what we wanted.

The MC33363B is a voltage-mode controller combined with a pulse-by-pulse current limit detection. In our application we wire pin 10 (Feedback) to the ground in order to operate the IC at the maximum duty-cycle (50%). Then every output pulse is interrupted when the primary current reaches the maximum limit. The nice feature is the flexibility to control the peak current and the oscillation frequency. As a matter of fact, you can easily adjust the peak current to account for the magnetostriction effects of your transformer. Any transformer used on an hysteretic SMPS will generate audible noise at low

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frequency. You either buy a good transformer whose price blasts your final target or, as offered by figure 1's circuit, you ensure a low the peak current. This prevents your transformer to "sing" during hysteretic operation. If you reduce the peak current you must either rise the primary inductance or the switching frequency (C9) to keep the output power constant.

The transformer features a primary inductance of 6.5mH

and respective auxiliary and power winding ratios of 0.2 and 0.06. It is available from Orega in the U.S. (San-Diego, CA). Thanks to a low leakage inductance, a simple RC network needs to be wired from drain to source. It does not significantly degrade the efficiency.

The measurement results are summarized in the below array (Tamb = 25°C):

<i>DC input level</i>	<i>Input Power</i>	<i>Output power</i>	<i>η</i>
120	34.8mW	0	–
120	565mW	332mW	58.8%
120	1.305W	880mW	67.5%
120	2.472W	1.677W	67.8%
325	42.2mW	0	–
325	549mW	337mW	61.4%
325	1.35W	893mW	65.9%
325	2.7W	1.68W	62.5%
369	44mW	0	–
369	568mW	337mW	59.7%
369	1.38W	880mW	63.5%
369	2.76W	1.65W	59.7%

MC34063 and IGBT reach 43mW offline standby power

In nowadays environment, the chase is open to trap every bit of wasted power. In that area, Switch-Mode Power Supplies (SMPS) bring the benefit of a better efficiency than linear supplies. However, if the efficiency is good at nominal load, it tends to degrade seriously when going down to small loads. Another concern from SMPS is their difficulty to deal with no-load conditions. As a natural reaction to the lack of load, the controller dictates its smallest duty-cycle but the commutation losses associated with the controller consumption remain prominent and the power drawn is usually well above the watt.

A nice way to build a low standby power SMPS is to implement the well known hysteretic architecture. The way the

supply works is very simple: if the output voltage is below the target, the controller delivers high-frequency pulses until the target is reached. When this condition occurs, the controller stops the pulse emission and the switch is kept fully opened until the output voltage drops below the threshold. Then a new train of pulses takes place. At no load, the controller thus issues packet of pulses to compensate the self-discharge of the output capacitor. As a matter of fact, only very few energy is necessary to compensate for it and the repetition rate of the pulse packet is very long. You obviously consume little power from the source.

Figure 2 depicts the implementation of the circuit using the MC34063 from MOTOROLA (Phoenix, AZ). This circuit include an 1.5A bipolar switch but we will simply use it as a follower.

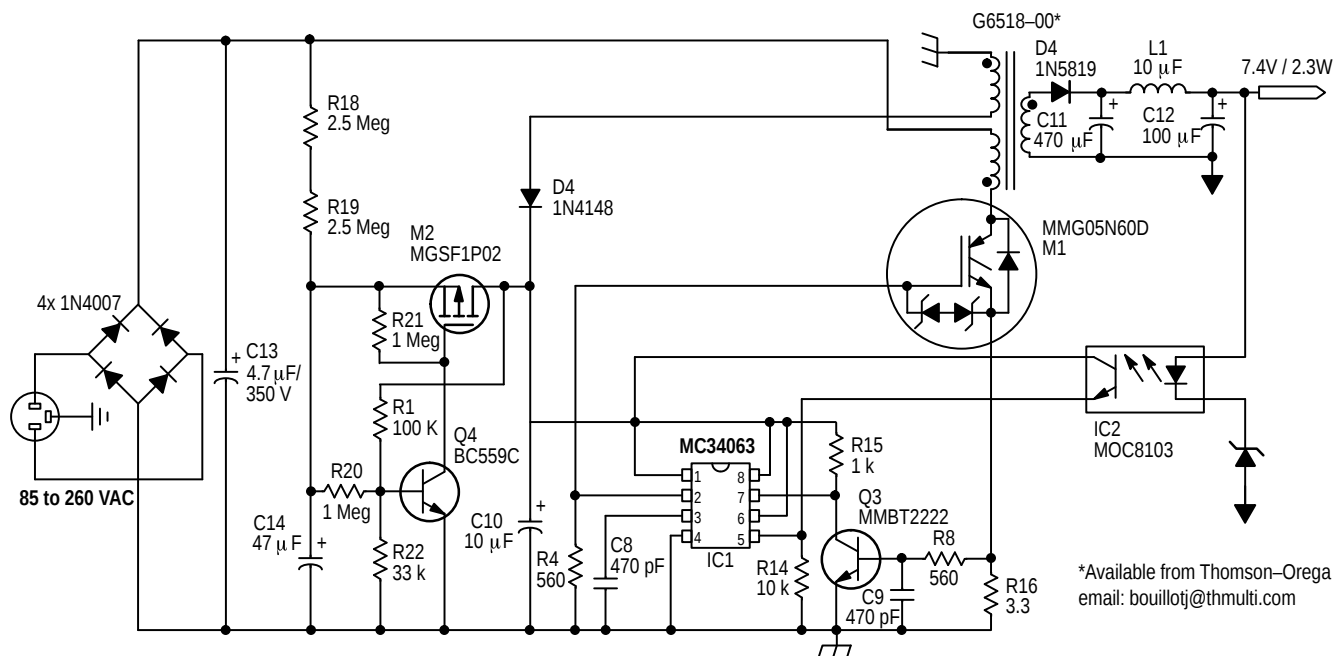


Figure 2. A MC34063 associated with an IGBT consumes less than 100mW on universal mains.

The device operates in current-mode thanks to Q3 which trips the IC when the sense resistor R16 drops around 550mV ($I_{peak} = 160\text{mA}$ in our example). As a matter of fact, you can easily tweak the peak current to account for the magnetostriction effects of transformers. Any transformer used on an hysteretic SMPS will produce audible noise at low frequency. Either you buy a costly transformer where construction precautions were taken to ensure low audible noise or, as offered by figure 1's circuit, you reduce the peak current. If you reduce the peak current you need to either rise the primary inductance or the switching frequency (C8) to keep the output power constant.

By using the new MMG05N60D IGBT, you naturally benefit from low parasitic capacitors which degrades the efficiency at low peak currents. When you open the high-voltage switch in a FLYBACK converter, the peak current does not drop to zero

immediately. Depending on the amount of COSS capacitance (CDG + CDS) the current keeps circulating while the drain rises: the component dissipates and the supply cannot reach a good efficiency. **Figure 2a** illustrates this poor behavior with an integrated high-voltage switcher (lateral MOSFET). At the opposite, the MMG05N60D features very low parasitic capacitors. As an example, the total gate charge to reach $V_{GS} = 10\text{V}$ is only 4nC! **Figure 2b** depicts the current taken from figure 1's circuit when operating at null output power. The little current tail is typical from the IGBT architecture. However, the MMG05N60D features an internal path to evacuate the trapped carriers at turn-off. The typical losses are around 6.5μJ for a $1\text{kV}/\mu\text{s}$ $dV_{CE}/dt@I_C = 300\text{mA}$. Finally, thanks to its accidental avalanche capability, you do not need to install an external voltage clipping network.

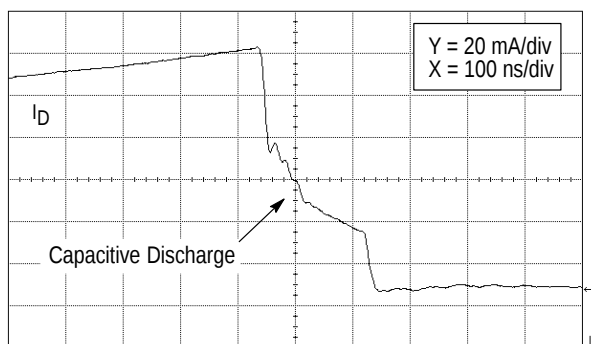


Figure 2a. Typical turn-off current waveform of a lateral MOSFET

Since the MC34063 exhibits an operating current typical of the mA, we cannot make it start the usual way. Keep in mind that we are chasing the mWs so a few k start-up resistor is prohibited! We have implemented a circuit made of Q4 and M2 that has already been published as a Design Idea by the author. When you apply the mains, C14 level starts to elevate.

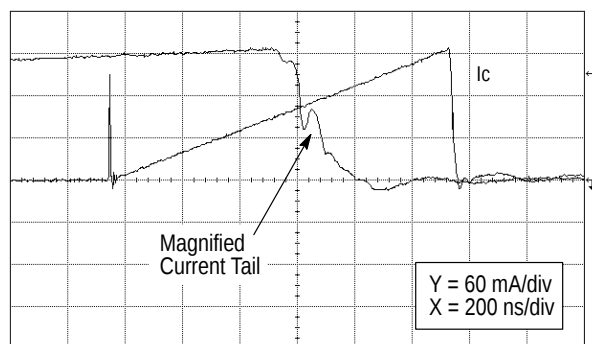


Figure 2b. Turn-off shot of the MMG05N60D collector current

Because of R20/R22 ratio, Q4 and M2 are kept open: MC34063 receives no power and the current drawn from C14 is very small. When sufficient level is reached, Q4 enters in conduction and pulls M2's gate toward the ground. At this time, its drain level goes up and thanks to R1, it strengthens Q4 conduction. C14 is thus brutally discharged into C10 and the

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MC34063 oscillates. D4 and the auxiliary level now take over and the circuit is self-supplied. With figure 1's values, you start with a 5M resistor. Of course, the charging time is rather long (1mn @ $V_{in} = 120V$) but applications like battery chargers do not care about it. As an alternative, one could imagine a big power supply (e.g. 200W for a TV) which first starts and gives the starting pulse to the MC34063 circuit. When the user

forces the standby mode via the remote control, it naturally shuts the big SMPS down, but the MC34063 circuit is still alive, consuming very little power to supply the μ Processor and the front panel L.E.D.

The measurement results are summarized in the below array ($T_{amb} = 25^{\circ}C$):

<i>DC input level</i>	<i>Input Power</i>	<i>Output power</i>	<i>η</i>
120	42.5mW	0	–
120	552mW	363mW	65.7%
120	1.14W	796mW	69.7%
120	3.25W	2.38W	73%
325	65mW	0	–
325	585mW	364mW	62.2%
325	1.17W	796mW	68%
325	3.38W	2.38W	70.7%
360	82.8mW	0	–
360	612mW	364mW	59.4%
360	1.224W	796mW	65%
360	3.52W	2.37W	67.3%

As you can see, the average input power is kept below 83mW at $V_{AC} = 250V_{eff}$, which is a good performance compared to figure 1's simplicity. The efficiency is also maintained at a good performance level even when the output power decreases.

Low-voltage bipolars and UC3845 boast low startup current

Popular UC384X-based Offline Switch Mode Power Supplies (SMPS) are usually self-supplied by an auxiliary winding. This winding feeds the main Pulse Width Modulator (PWM) controller once the steady-state operation takes place. Despite nowadays technologies, the controller always needs a small startup current which clearly plagues the efficiency in very low-power standby SMPS ($P_{out} = 500mW$). Any wasted source (startup network, controller supply etc.) will add to the power drawn from the mains and significantly degrades the overall efficiency. In these conditions, obtaining an acceptable efficiency at high-line represents a tough design task. Some tricks can totally stop the startup current,

e.g. by using a high-voltage bipolar or MOSFET (reference [1]). However, the addition of a high-voltage component burdens the bill-of-material cost of a small-power SMPS.

Figure 3 represents a cheap solution where the startup current can be programmed at any value, whatever the PWM IC. Implemented around two *low-voltage* off-the-shelf bipolars, the circuit brutally connects the controller to capacitor C6 when its voltage level is adequate. This level determines the capacitor C.V product which shall be large enough to feed the PWM IC until the steady-state operation takes over. The designer can then select any small startup current to charge the bulk capacitor, in agreement with his time-constant recommendations. During the charge, series PNP Q1 is locked-off and prevents any current consumption from the main controller. When a defined level is reached, Q2 starts to pull Q1's base to ground. Its collector voltage rises and further saturates Q2 through the reaction resistor R7. The charged bulk capacitor is now fully connected to the PWM IC who starts to oscillate. A few cycles later, the auxiliary winding takes over and Q1 stays on.

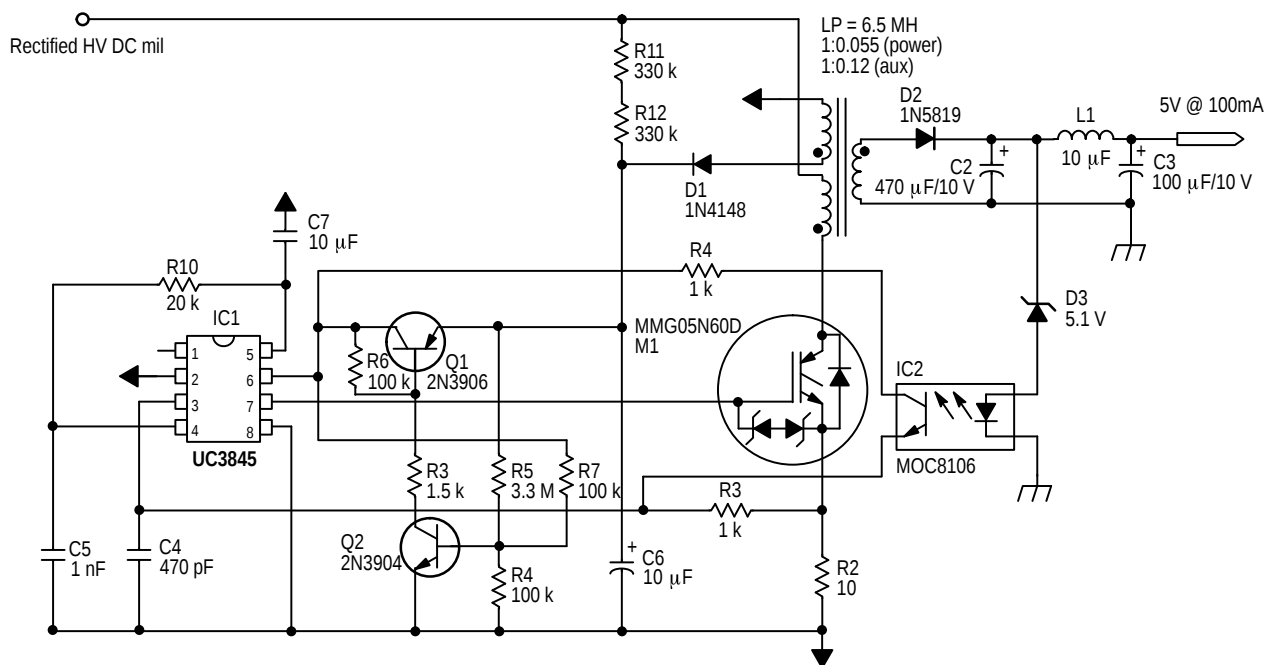


Figure 3. A pair of bipolars turns any PWM controller into a low-current startup device

The level $V_{threshold}$ at which Q2 turns on is simply given by: $V_{BEQ1} \cdot \frac{R4 + R5}{R4}$. If we consider a linear charge from the startup network $R11+R12$, then the time at which Q2 starts to conduct is given by:

$\frac{C6 \cdot V_{threshold}}{I_{startup}}$, with $I_{startup} \leq \frac{V_{DCrail}}{R11 + R12}$. With the given values, C6 feeds the IC when its voltage reaches 17V in less than 350ms @ VDC = 350V.

Figure 1's circuit offers better than 50% efficiency at high-line (250VAC), drawing less than 1W for a 500mW output power. The new MOTOROLA (Toulouse, France) IGBT MMG05N60D also participates to the performance thanks to its low capacitive parasitic elements (5nC gate charge!). The commutation losses are thus kept at a minimum level.

Figure 4 details the startup phase of the low-power SMPS shot at very low-line (70VAC!).

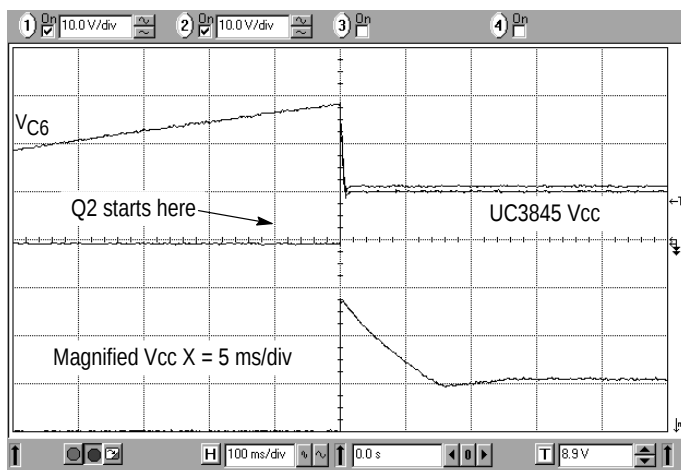


Figure 4. Low-line startup sequence of the two-bipolar network

Reference 1: Christophe BASSO, "Low-cost MOSFET quashes power resistor", EDN Design Idea June 6, 94.

Low-cost IGBT charges batteries from universal mains

New power supplies for portable equipments need to be light and provide the international traveler with a convenient universal input. If Switch Mode Power Supplies (SMPS) naturally benefits from this universal input, they have difficulty to compete in price with standard linear solutions based on a metallic transformer. Thanks to the introduction of new IGBTs

from MOTOROLA, **figure 5** offers a cheap solution for charging a battery from the AC line. The MMG05N60D sustains up to 600V with an avalanche capability comparable to that of a MOSFET of same characteristics. However, thanks to its small die area, the MMG05N60D will advantageously replace the standard high-voltage MOSFET selected for output powers up to 8W. Furthermore, its SOT-223 package is fully pin and size compatible with the DPAK standard.

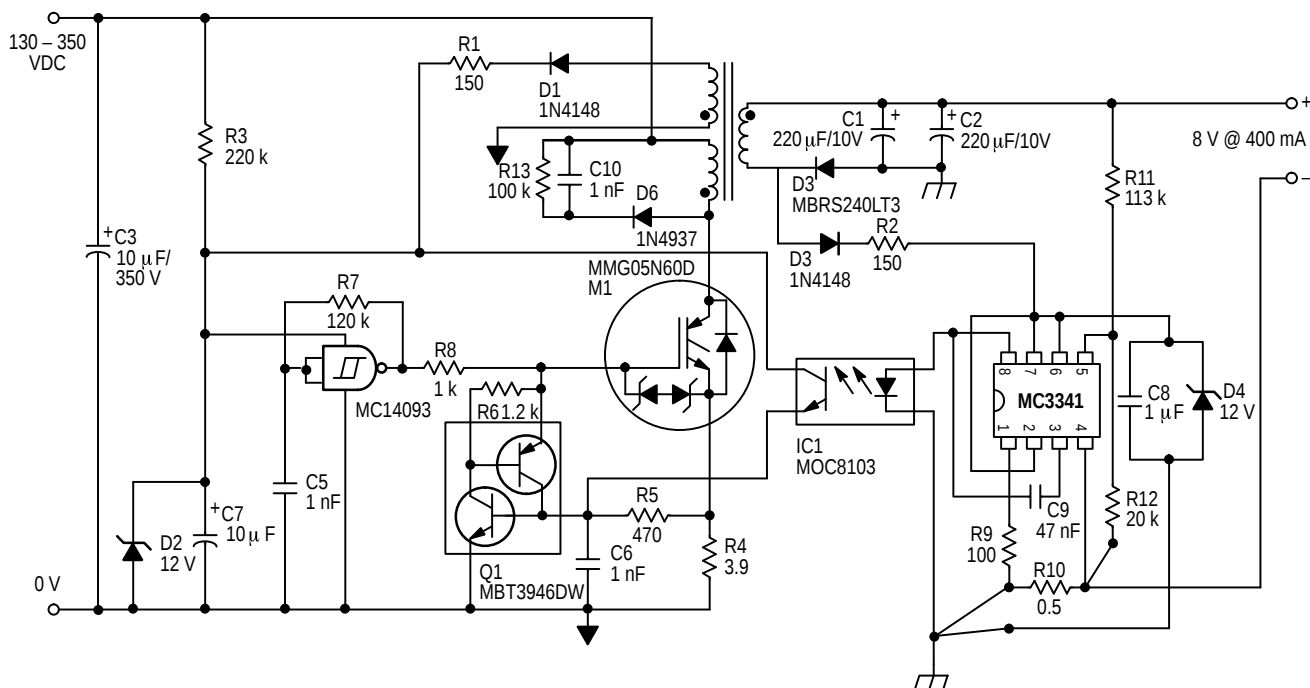


Figure 5. This offline supply charges a battery with a square I/V shape.

The offline supply depicted by figure 5 implements a current-mode technique where the latch is provided through a discrete thyristor (Q1, double bipolar). The 50% duty-cycle clock can be generated with off-the-shelf components as long as they are able to generate a 50kHz square-wave with a low start-up current (fixed by R3). We selected a NAND Schmitt trigger extracted from a classical MC14093. Thanks to its very low gate charge (7nC), you can drive the MMG05N60D by simple logic gates and does not require any specific driver. In the application, the IGBT receives its bias

voltage through R8, a 1k resistor. This rather high value is needed in order to not disturb the oscillator when the thyristor pulls the gate to ground. As we said, the low gate-charge authorizes the use of such elevated drive impedance. When the IGBT is turn on, the primary current starts to rise as the voltage does across R4. When the peak current is reached, or when 700mV appear upon C6, Q1 turns on and stops the IGBT conduction. As for any flyback, the energy transfer takes place and charges C1-C2, the output capacitors.

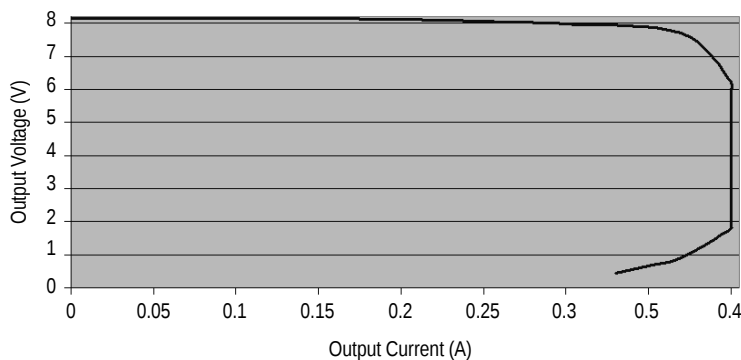


Figure 6. Square I/V characteristic imposed by the MC33341

In these low-cost structures, you can easily control the output voltage/current by offsetting the DC level present across C6 connections. This is done through an optocoupler driven by a dedicated battery charger circuit, the MC33341. This component includes a dual control loop which regulates either the current (sensed by R10) or the voltage (sensed by R11/R12). When the current is under limit, the MC33341 regulates the output voltage at its nominal value (e.g. 8V) and lets the output current grow. When this output current reaches the internal threshold ($0.2V/R10$) the current loop takes over and leads the optocoupler to transform the supply into a constant current source. **Figure 6** portrays this typical output shape measured on the supply. A SPICE model of the

MC33341 is available for download at:
http://mot2.indirect.com/models/bin/batmag_ic.html

As you will note, the dot marked terminals of the transformer are not wired in the usual way. This comes from the fact that the output can be totally shorted when you connect an empty battery. If you short the output, the flyback auxiliary winding decreases but the MC33341 supply disappears, leaving the supply without a current brake. To avoid this situation, the output transformer is connected so as to benefit from the flyback voltage but also from the forward voltage: when the flyback voltage is constrained to a low level by the output, the forward voltage is present and powers the MC33341. The auxiliary winding also benefits from this structure.

The supply has been tested with a transformer exhibiting a primary inductance of 6mH and respective auxiliary and power winding ratios of 0.12 and 0.06. It is important that you prevent the supply entering the continuous mode in order to cancel any turn-on losses. The MMG05N60D has been designed to keep the current tail losses at 6 μ J typically (@ $I_c=0.3A$, $T_j=125^\circ C$ and $dV_{CE}/dt = 1kV/\mu s$).

IGBT powers ultra low-cost offline wall adapter

Universal AC to DC wall adapters need to be light and compact in order to meet the international traveler

requirements. In that area, Switch Mode Power Supplies (SMPS), are the natural candidate for these applications. However, the total circuitry cost shall be kept low in order to compete with actual metallic transformer solutions. **Figure 6's** schematic implements the new MMG05N60D IGBT recently introduced by MOTOROLA Semiconductors and represents a solution where space and cost are at premium. Associated with a dual bipolar and a few surrounding components, the supply delivers up to 6W with an efficiency better than 70% @ $V_{in}=120VAC$.

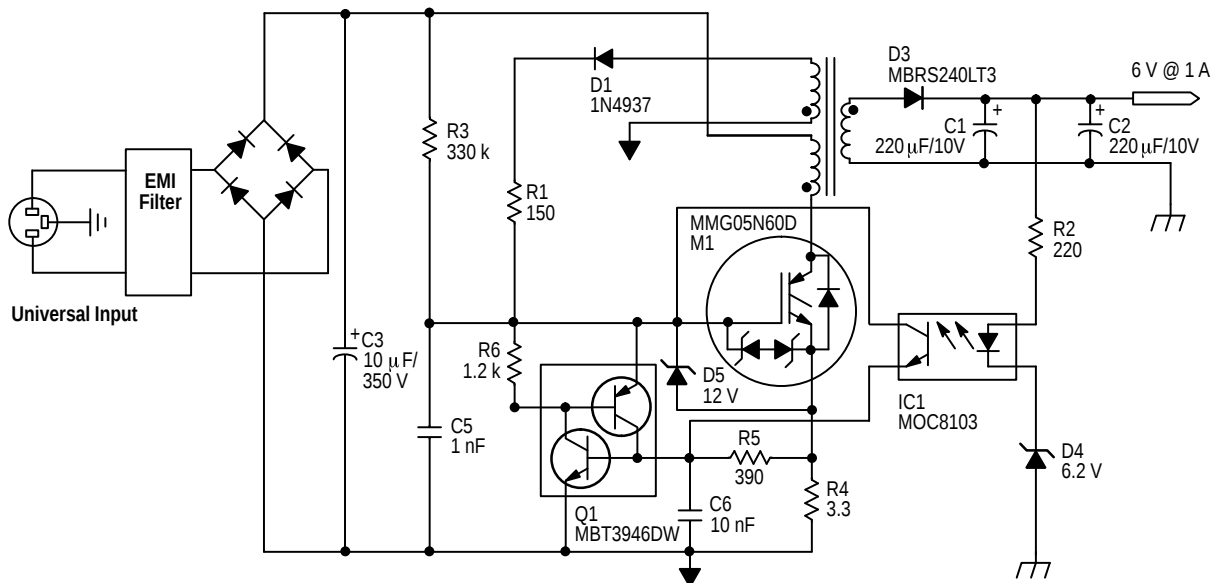


Figure 7. This SMPS delivers 6W with a bunch of semiconductors surrounding a fast IGBT

At power-on, R3 charges C5 and the input capacitance of the IGBT (5nC gate charge!). When the threshold voltage is reached, M1 starts to close and applies the rectified DC rail across the primary. A voltage immediately appears on D1's

anode and accelerates the turn-on phase of the IGBT. This speed-up pulse allows a fast turn-on of the IGBT without degrading the overall efficiency. The collector current rises with a slope of

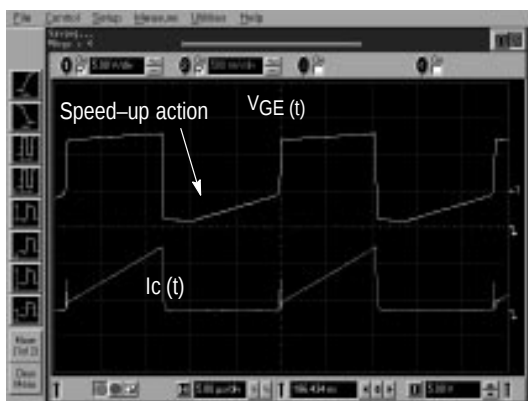


Figure 8. Operation details of the IGBT SMPS

V_{in}/L_p and once the bipolar thyristor threshold is reached ($I_p = 0.7V/R_4$), the gate is pulled down to ground. The FLYBACK transfer takes place, feeding the output capacitor and elevating the output voltage. Regulation is obtained by offsetting the thyristor threshold and interrupting the primary current sooner. We adopted a very cheap regulation network but a complete square I/V battery charger can easily be implemented using the MC33341 in place of zener D4. **Figure 8** details the operating waveforms.

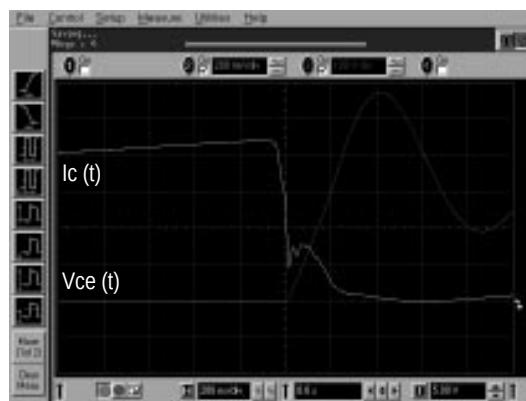


Figure 9. Turn-off losses are kept low thanks to a novel design

The supply has been built with an off-the-shelf transformer featuring an L_p of 6mH and respective auxiliary and power winding ratios of 0.12 and 0.06. It is important to keep the supply operating in discontinuous mode in order to cancel any turn-on losses. The MMG05N60D has been carefully designed to keep the current tail losses at 6 μ J typically (@ $I_c=0.3A$, $T_j=125^\circ C$ and $dV_{CE}/dt = 1kV/\mu s$) as **figure 9** portrays. Furthermore, it includes a free-wheel diode which makes the device avalanche proof. The MMG05N60D

represents a cost effective solution versus a high-voltage MOSFET in offline SMPS up to 10W.

A method to understand the way this kind of supply works is to dive into the nitty-gritty of the system with the help of a

SPICE simulator. We simulated figure 1's schematic with INTUSOFT's IsSpice4 (San-Pedro, CA) and the main results are gathered in **figure 10**.

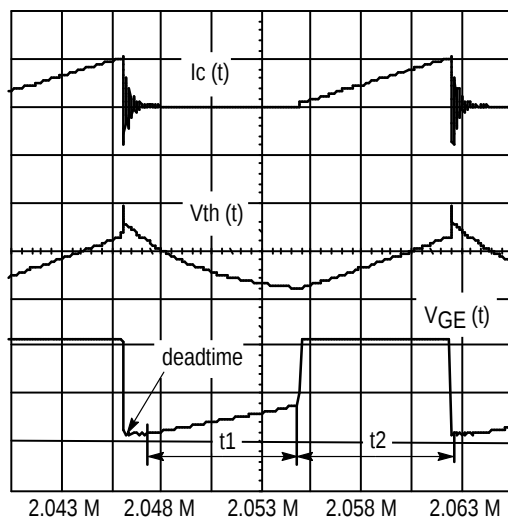


Figure 10. IsSpice4 simulation results of the IGBT power supply

As one can see, they nicely agree with the real measurements, highlighting the effects of the speed-up pulses. t_1 is the time needed by the VGE voltage to reach the threshold level of 3.5V. $I_c(t)$ ramps from 0 to I_p and its slope is defined by V_{in}/L_p . The deadtime corresponds to the thyristor reset and is constant at around 3 μ s. Summing up these times leads to the operating period. The IGBT SPICE model can be obtained by e-mail upon request to the author.

IGBT and UC3845 combine into low-cost offline hiccup supply

In equipments requiring a stand-by mode, e.g. VCRs, TVs

etc, you must still supply the microprocessor when everything is asleep in order to receive and interpret the wake-up signal from the remote control. The power consumption of the system is rather low and classical Switch-Mode Power Supplies chips represent a clear overkill for a 100–200mW output level. The active solution also needs to be cost effective compared to the standard structure using a metallic transformer.

Figure 11 schematic shows how an UC3845 can simply drive the new MMG05N60D 600V IGBT recently introduced by MOTOROLA (Toulouse, France).

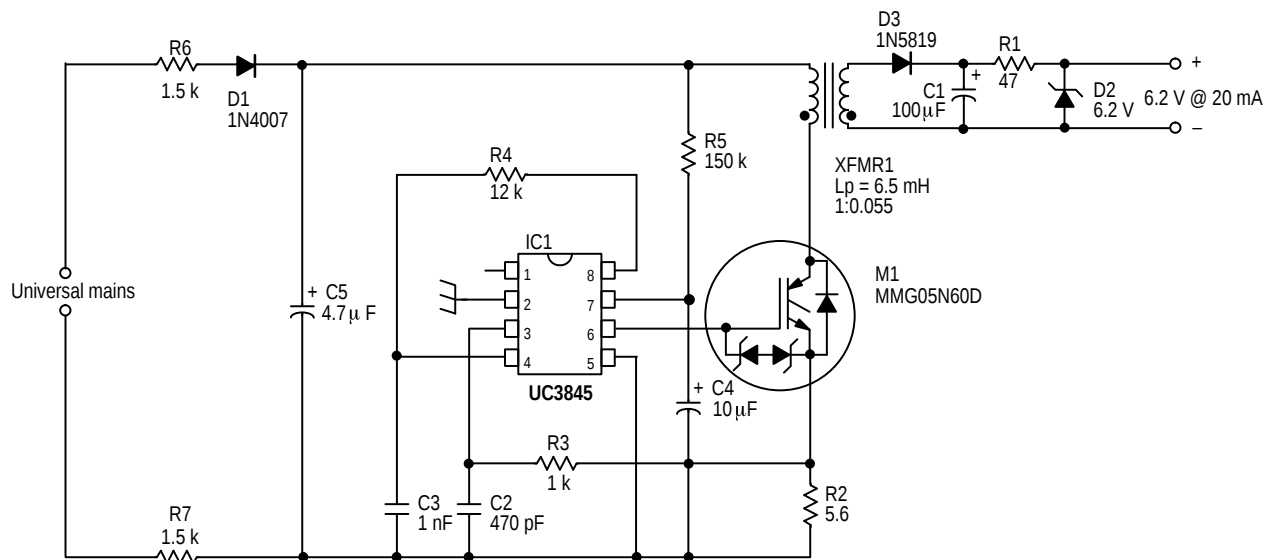


Figure 11. A simple UC3845 and the MMG05N60D build an economical stand-by SMPS

In this application, the 3845's functionality is reduced to its maximum by grounding the internal error amplifier via the feedback pin. In that case, only the clock generator and the peak current detector remain active. At power-on, R3 pulls-up C4 potential above the 3845's Under Voltage

LockOut level (UVLO). The start-up current does not exceed a few hundred of μ A and consequently does not waste any power. When IC1 enters operation, pin 6 delivers a driving signal to M1. Thanks to its insignificant gate charge (5nC!), it does not

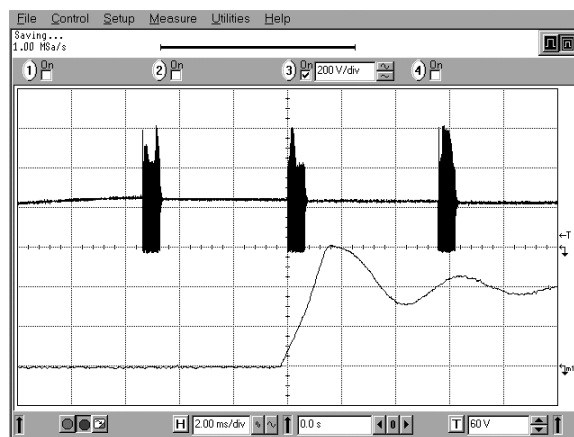


Figure 12. The MMG05N60D safely clamps the collector voltage around 600V

excessively hampers C4 voltage and a burst of pulses can take place. Power is transferred to the output when M1 turns off as in classical FLYBACK mode. The UC3845's operating current discharges C4 until the supply falls under the UVLO level: IC1 stops operating and enters start-up mode again. An hiccup operation takes place whose recurrence is defined by IC1's operating current, R5 and C4. The peak current is regulated by the UC3845 and allows operation over the universal mains. Since the circuit operates at constant output power, the peak current can be extracted by the following

formula:
$$I_p = \sqrt{\frac{2 \cdot P_{out}}{L_p \cdot F_{osc}}}$$
 With an internal error amplifier

clipped at 1V, R2 is simply $1/I_p$.

Thanks to its avalanche capability, the MMG05N60D does not require any clipping network, as **figure 12** unveils by magnifying a burst. The shot was taken with a 250VAC line. The IGBT case temperature was less than 40°C at an ambient of 25°C.

The transformer XFMR1 is typically an off-the-shelf model as long as the supply operates in discontinuous mode. It is the case with the values given upon **figure 12's** drawing.

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