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MC44603 in a 110 W Output SMPS Application (80-140 Vrms and 180-280 Vrms Mains Voltages)

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The purpose of this application note is to present a way of designing the MC44603 in a fly–back dedicated to a 110 W output power supply in two distinct cases: 110 Vrms mains and 220 Vrms mains.

For this range of power, the discontinuous mode must be chosen as it limits the stress on the power switch and the output diodes. This kind of working can be guaranteed, thanks to the demagnetization arrangement of the MC44603.

This application note considers both high and low mains voltages.

— In the high voltage a.c. line case, it deals with both MOSFET and BIPOLAR transistor use.

— In the low mains voltage case, only the MOSFET solution is considered as the inductor peak current is high.

The MC44603 is a high flexibility SMPS controller. It offers a wide variety of protection (accurate maximum current limitation, Vcc overvoltage detection, fold–back, demagnetization, soft start).

In addition, distinct working modes can be used with this circuit:

— **a fixed frequency working mode**
Here, the frequency is given by the circuit oscillator.

— **a variable frequency working mode**
This kind of working can be obtained when the chosen oscillator period is shorter than the cycle time for transformer magnetization and complete demagnetization.

— **a stand–by mode**
This reduces the working frequency during this phase. This method reduces the stand–by losses that are mainly proportional to the switching frequency, and avoids the unstable very high frequency working that is generally associated with traditional free frequency systems.

In the MC44603 design proposed here, the variable frequency mode is not used. Indeed, as shown later, the fixed frequency mode enables a more accurate control of the power that is drawn from the mains.

I — THE APPLICATIONS

OUTPUT: 110 W

In the following calculations, the maximum input power will be considered equal to 135 W ($\eta \approx 80\%$).

Application 1: 110 V mains 80 Vrms ↔ 140 Vrms

Application 2: European mains 180 Vrms ↔ 280 Vrms

The outputs:	120 V	0.5 A
	28 V	1.0 A
	15 V	1.0 A
	8 V	1.0 A

Consequences for the Transformer

This output voltage values are obtained using four secondary windings of the transformer. Now, as a diode is located between the output capacitor and the winding, this diode voltage must be taken into account in the choice of the number of turns of each of the windings.

So, if the diode voltage is considered equal to 1 V, the desired voltages on the windings are nearly:

121 V
29 V
16 V
9 V

To obtain the output voltage values of the specification with a good accuracy, we need at least 3 turns for the 9 V and then:

9 V	3 turns
16 V	5 turns
29 V	10 turns
121 V	40 turns

This solution seems to be correct, since with three turns, the 9 V should be well coupled; also, 40 turns (for the 120 V) is not too large a value (no ferrite saturation).

In our application, in order to obtain a very well regulated high voltage output, an opto coupler is used. By this means, the regulation uses feedback drawn from the 120 V output.

II — GENERAL DESIGN EQUATIONS

Calculation of the Main SMPS Parameters

II–1 — Fixed frequency working mode:

To obtain a fixed frequency discontinuous working mode (the discontinuous mode is guaranteed by the demagnetization section), the transformer primary inductor magnetization and demagnetization cycle must be shorter than the oscillator period (that is the chosen working period) for any working point.

This condition can be expressed by the following inequality:

$$\mathrm{Ton} + \mathrm{Toff} \le \mathrm{Tosc} \qquad \text{(ineq 1)}$$

where: Ton is the on–time, Toff is the off–time

Tosc is the MC44603 oscillator period

The current increases linearly during the on–time and decreases linearly in the inductor for the off–time.

So, $$Ipk = \frac{Vin}{L} \times Ton \quad \text{(eqn 1)}$$

and, $$Ipk = \frac{NVo}{L} \times Toff \quad \text{(eqn 2)}$$

where: Ipk is the primary inductor peak current

Vin is the rectified a.c. line voltage (fly–back input voltage)

L is the primary inductor value

NVo is the output voltage considered in the primary side

(N: turn ratio)

So, $$Ton + Toff = L \times Ipk \times \frac{Vin + NVo}{Vin \times NVo} \quad \text{(eqn 3)}$$

In addition, the energy drawn through the transformer during one cycle is:

$$E = \frac{1}{2} \times L \times Ipk^2 \quad \text{(eqn 4)}$$

So, Pin (input power) is:

$$Pin = \frac{1}{2} \times L \times Ipk^2 \times fosc \quad \text{(eqn 5)}$$

where fosc is the MC44603 oscillator frequency

Using equations 3 and 5 and inequality 1,

condition required to work in a fixed frequency mode:

$$L \times fosc \le \frac{1}{2 \times Pin} \times \left(\frac{Vin \times NVo}{Vin + NVo}\right)^2 \quad \text{(ineq 2)}$$

II–2 — Peak inductor current expression

Pin is given by the following expression:

$$Pin = Vin \times Iin \quad \text{(eqn 6)}$$

where Iin is the input current

and $$Iin = \frac{1}{2} \times Ipk \times d \quad \text{(eqn 7)}$$

where d is the duty cycle

Now, $$Ton = d \times Tosc \quad \text{(eqn 8)}$$

Using equations 1, 7 and 8:

$$Iin = \frac{1}{2} \times L \times Ipk^2 \times \frac{fosc}{Vin} \quad \text{(eqn 9)}$$

Using equations 6 and 9:

$$Ipk = \sqrt{\frac{2 \times Pin}{L \times fosc}} \quad \text{(eqn 10)}$$

Thus, Peak inductor current:

$$(Ipk)max = \sqrt{\frac{2 \times (Pin)max}{L \times fosc}} \quad \text{(eqn 11)}$$

II–3 — Power switch on–time losses

II–3–1 — MOSFET:

Rdson being the MOSFET on–time resistor, the on–time losses can be calculated from:

$$Pon = \left(\frac{1}{Tosc}\right) \times \int_0^{Ton} Rdson \times I_T{}^2 \times dt \quad \text{(eqn 12)}$$

where I_T is the MOSFET current.

So, as: $$I_T = Vin \times \frac{t}{L} \quad \text{(eqn 13)}$$

$$Pon = \frac{1}{3} \times Rdson \times Vin^2 \times \frac{Ton^3}{(Tosc \times L^2)} \quad \text{(eqn 14)}$$

Consequently, using equations 1, 5 and 14, the following expression can be written:

$$Pon = \frac{2 \times \sqrt{2}}{3} \times Rdson \times \frac{Pin}{Vin} \times \sqrt{\frac{Pin}{L \times fosc}} \quad \text{(eqn 15)}$$

Consequently,

maximum MOSFET on–time losses

(eqn 16)

$$(Pon)max = \frac{2 \times \sqrt{2}}{3} \times Rdson \times \frac{(Pin)max}{(Vin)min} \times \sqrt{\frac{(Pin)max}{L \times fosc}}$$

Using the (Ipk)max value, the following equation could also be written:

$$(Pon)max = \frac{1}{3} \times Rdson \times (Ipk)max^2 \times dmax$$

where dmax is the maximum duty cycle.

II–3–2 — BIPOLAR TRANSISTOR

$$Pon = \frac{1}{Tosc} \times \int_0^{Ton} V_{CE} \times I_T \times dt \quad \text{(eqn 17)}$$

So: $$Pon = V_{CE} \times \langle I_T \rangle \quad \text{(eqn 18)}$$

Thus: $$Pon = V_{CE} \times Iin \quad \text{(eqn 19)}$$

where Iin is the input current

So: $$Pon = V_{CE} \times \frac{Pin}{Vin} \quad \text{(eqn 20)}$$

Consequently,

maximum BIPOLAR transistor on–time losses

$$(Pon)max = V_{CE} \times \frac{(Pin)max}{(Vin)min} \quad \text{(eqn 21)}$$

II–4 — Maximum duty cycle

The duty cycle, d, is equal to: $$d = \frac{Ton}{Tosc} \quad \text{(eqn 22)}$$

Now, $$\mathrm{Ton} = L \times \frac{\mathrm{Ipk}}{\mathrm{Vin}} \quad \text{(eqn 23)}$$

so, $$d = L \times \mathrm{fosc} \times \frac{\mathrm{Ipk}}{\mathrm{Vin}} \quad \text{(eqn 24)}$$

Thus, using equation 10:

$$d = \sqrt{2 \times \mathrm{Pin} \times L \times \frac{\mathrm{fosc}}{\mathrm{Vin}^2}} \quad \text{(eqn 25)}$$

So, ***maximum duty cycle:***

$$\mathrm{dmax} = \sqrt{2 \times (\mathrm{Pin})\mathrm{max} \times L \times \frac{\mathrm{fosc}}{(\mathrm{Vin})\mathrm{min}^2}} \quad \text{(eqn 26)}$$

II–5 — Other design parameters

The other parameters that must be taken into account are:

(V_T)max: the maximum voltage the power switch must face:

$$(V_T)\mathrm{max} = (\sqrt{2} \times (\mathrm{Vin})\mathrm{max}) + (N \times 120) \;\; (V) \quad \text{(eqn 27)}$$

where: (Vin)max is the maximum rms a.c. line voltage (140 V or 280 V according to the line)

120 is the maximum output voltage

N is the transformer turn ratio between the 120 V output winding and the primary inductor

(V_D)max: the maximum voltage the 120 V output diode must face:

$$(V_D)\mathrm{max} = \left(\sqrt{2} \times \frac{(\mathrm{Vin})\mathrm{max}}{N}\right) + 120 \;\; (V) \quad \text{(eqn 28)}$$

(ni): the transformer ferrite saturation parameter

To avoid any risk of saturation in the transformer, the inductor peak current must be lower than (ni/ n_p), where n_p is the turns number of the primary inductor.

Now, $$n_p = N \times n_{120V} \quad \text{(eqn 29)}$$

where n_{120V} is the turns number of the 120 V winding

So, $$ni = N \times n_{120V} \times \mathrm{Ipk} \quad \text{(eqn 30)}$$

These parameters are the main elements that have to be taken into account as they allow you to choose the power switch, the diodes, the transformer and the working frequency.

As shown by the preceding calculations, the design parameters depend on some elements like N or the value of (fosc x L).

SUMMARY

Condition Required to Work in a Fixed Frequency Mode	$L \times \mathrm{fosc} \le \frac{1}{2 \times \mathrm{Pin}} \times \left(\frac{\mathrm{Vin} \times \mathrm{NVo}}{\mathrm{Vin} + \mathrm{NVo}}\right)^2$
Maximum Peak Inductor Current	$(\mathrm{Ipk})\mathrm{max} = \sqrt{\frac{2 \times (\mathrm{Pin})\mathrm{max}}{L \times \mathrm{fosc}}}$
Maximum Power Mosfet On–Time Losses	$(\mathrm{Pon})\mathrm{max} = \frac{1}{3} \times \mathrm{Rdson} \times (\mathrm{Ipk})\mathrm{max}^2 \times \mathrm{dmax}$
Maximum Bipolar Transistor On–Time Losses	$(\mathrm{Pon})\mathrm{max} = V_{CE} \times \frac{(\mathrm{Pin})\mathrm{max}}{(\mathrm{Vin})\mathrm{min}}$
Maximum Duty Cycle	$\mathrm{dmax} = \sqrt{2 \times (\mathrm{Pin})\mathrm{max} \times L \times \frac{\mathrm{fosc}}{(\mathrm{Vin})\mathrm{min}^2}}$

III — APPLICATION 1: 110 V INPUT

III–1 — Choice of the transformer

One way to use the above design equations, consists of drawing up a table showing how the main SMPS parameters vary with the value of the turn ratio.

To calculate these values, it is necessary to know the input power level. This value is taken equal to (135 W) in our application (135 W corresponds to an efficiency equal to about 80%. The application results will show that this assumption ensures a desirable margin with the nominal input voltage).

On the other hand, the parameters calculation shows that (L x fosc)max is the (L x fosc) value that results in the lowest (Ipk)max and (P_{on})max ones (refer to Ipk or P_{on} expressions). This (L x fosc) value is the maximum one that guarantees a fixed frequency working for any working point (refer to section II–1). The SMPS parameters given in the following table are calculated using this threshold value.

Choice criteria and definition of the transformer:

As shown by the following table, the higher the turn ratio (N) is, the lower the peak current is. Now, the (ni)max is proportional to N and the voltage the transistor must face, increases when N rises. That is why an optimal N value must be chosen.

In fact, there are three main choice criteria:

— the peak current and the on–time losses.
N must be as large as possible in order to reduce the peak current and the on–time losses

— the voltage the power MOSFET must face.
Indeed, this voltage must be as low as possible to reduce its cost and in order to decrease the Rdson. That is why a MOSFET 400 V should be used. It is necessary to have a safety voltage margin, to avoid the need to incorporate a lossy and costly clamping network that would cut the voltage spikes due to the leakage inductor (refer to Figure 1). Consequently, N must be chosen lower than 1.25 ($V_T$max = 350 V).

— the transformer must be well coupled.
This is to obtain a consistently accurate regulation of the output and to reduce the leakage inductor and hence the turning off spikes (refer to Figure 1). That is why a low air–gap ferrite must be used. Practically, a ferrite whose (ni) is lower than 200 A.turns, seems to be a good choice. So, N must be lower than 1. (The appendix gives details of OREGA transformers; the SMT4 suits our application).

Finally, in order to use a ferrite (ni = 180, A_L = 250 nH/turns^2), (N = 0.75) seems to be a suitable value.

Indeed, this value should result in a well coupled transformer with a low leakage inductor value. On the other hand, if we do not take into account the turning off spikes, then the theoretical highest value the power MOSFET must face, is 290 V. Consequently, with a 400 V power switch, only a low loss clamping arrangement is required.

The chosen MOSFET, is the MTP10N40E (Rdson = 0.55 Ω, 400 V).

Consequently,

$$L_p = A_L \times (N \times 40)^2 \longrightarrow Lp \approx 225\ \mu H$$

and the optimal working frequency is:

$$L_p \times fosc \leq 9.3 \longrightarrow fosc \leq 41.3\ kHz$$

So, the following values can be chosen:

Lp = 225 µH
fosc ≈ 40 kHz (Rref = 10 kΩ, C_T = 1nF)

and then: Ipk ≈ 5.4A

N	(L.fosc)max	(Ipk)max (A)	(V_T)max (V)	(V_D)max (V)	MOSFET on losses/ Rdson (W/Ω)	(ni)max
0.50	5.6	6.9	260	520	5.7	139
0.75	9.3	5.4	290	390	4.3	162
0.90	11.0	5.0	300	340	4.1	180
1.00	12.5	4.6	320	320	3.7	184
1.25	14.9	4.3	350	280	3.5	215
1.50	17.3	4.0	380	250	3.2	240
2.00	21.9	3.5	440	220	2.8	281

NOTE: N: turn ratio (refer to II–5)
(V_T)max: maximum voltage the power switch must face
(V_D)max: maximum voltage the 120 V output diode must face

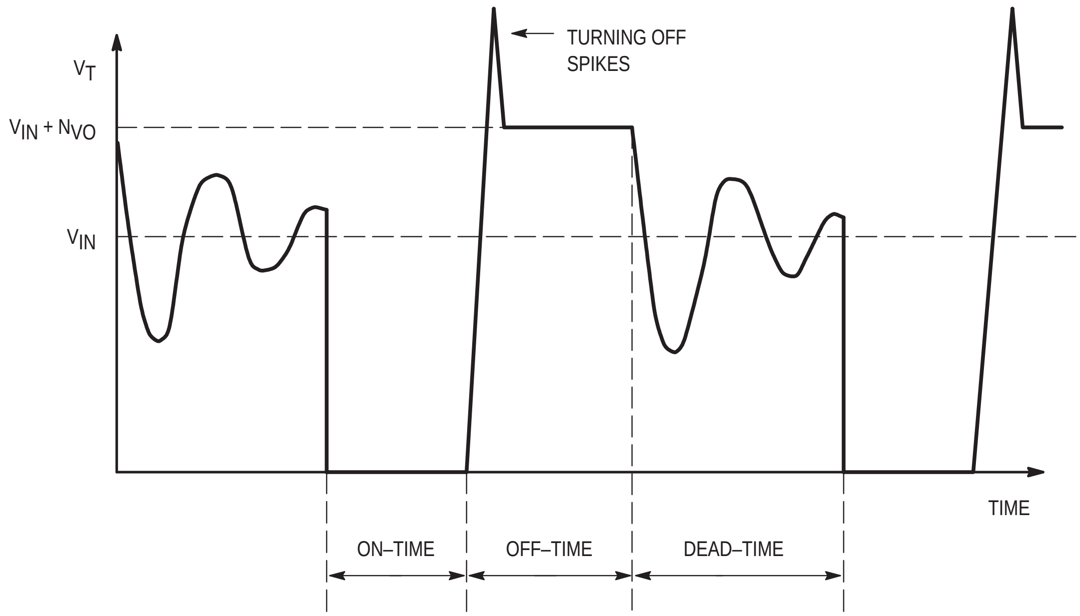


Figure 1. Voltage Spikes Due to the Leakage Inductor

III–2 — MC44603 pins use: (refer to the application schematics)

1 — Vcc (PIN 1):

The pin Vcc must be connected to a transformer auxilliary winding. This extra winding turns number can be taken equal to 5 in order to obtain a Vcc nearly equal to 15 V.

2 — Vc and OUTPUT (pins 2 and 3):

Vc is the output high state of the circuit. This pin offers the possibility of setting the output source current at a different level than the sink current but it is no use in our case.

In fact, a resistor of 33.2 Ω must be connected between the output and the MOSFET gate to make the switchings smoother. A resistor of about 1 kΩ can be connected between the gate and the ground (or the current sense external resistor) to avoid any inadvertent MOSFET switching on due to noise (see Figure 2).

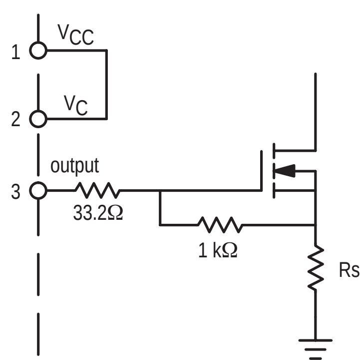


Figure 2.

3 — Foldback (pin 5):

Part of Vcc must be applied to this pin thanks to a resistor divider. This voltage value must be slightly higher than 1 V in normal use, so that this value drops below this threshold value as soon as an overload occurs.

4 — Overvoltage protection (pin 6):

This pin can remain free and then, the Vcc threshold level is fixed equal to nearly 17 V.

On the other hand, to make detection quicker and more accurate, an external resistor divider can be used with a diode and an integration capacitor (refer to the proposed application). The resistor divider is not directly connected to the Vcc because Vcc has a high time constant (refer to the application schematic — Figure 4).

5 — Current sense (pin 7):

The current sense resistor must be designed in order to limit the current below the maximum peak calculated in section II in order to limit the power that the converter is able to draw from the mains; in a fixed frequency mode, Pin = 1/2 x L x Ipk^2 x fosc.

Now, (Ipk)max = 5.4A

In addition, the (Vcs) clamp level is nearly 1 V (refer to the data sheet).

So, (Rs) the current sense resistor, must be equal to

(1 V / 5.4A), that is nearly: 0.18 Ω.

This value can be obtained using a 1W, 0.2 Ω resistor and a resistor divider (442 Ω, 3.16 kΩ) (refer to Figure 3).

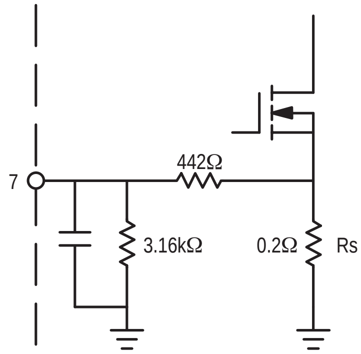


Figure 3.

Finally, as the fixed frequency mode is obtained for any working point, the peak current limitation results in an accurate input power limitation (135 W in this application note).

6 — Oscillator (pins 10 and 16):

The oscillator frequency is determined by the couple (C_T, Rref) (refer to the data sheet).

Even if capacitors have discrete values, the choice of Rref allows you to fix precisely the oscillator frequency (however, Rref also fixes the internal current source (Iref), which must be lower than 500 µA and higher than 100 µA).

$$\text{fosc} = 40\ \text{kHz} \longrightarrow \text{Rref} = 10\,\text{k}\Omega$$
$$\longrightarrow C_T = 1\text{nF}$$

7 — Stand–by mode (pins 12 and 15):

In the MC44603, it is possible to reduce the working frequency when little power is being drawn from the mains (stand–by mode).

This stand–by frequency is fixed by connecting a resistor R_{Fstby} to pin 15, while the power level at which the stand–by mode must be applied is determined by connecting another resistor R_{Pstby} to pin 12 (this power level is labelled P_{thL} in the data sheet).

In the data sheet, the equations needed to calculate R_{Fstby} and R_{Pstby} are indicated.

Using them, to obtain a power level equal to 10 W and a stand–by frequency equal to 20 kHz, the calculated R_{Fstby} and R_{Pstby} values are:

$$R_{Pstby} = 8.45\ \text{k}\Omega \text{ and } R_{Fstby} = 22.1\ \text{k}\Omega$$

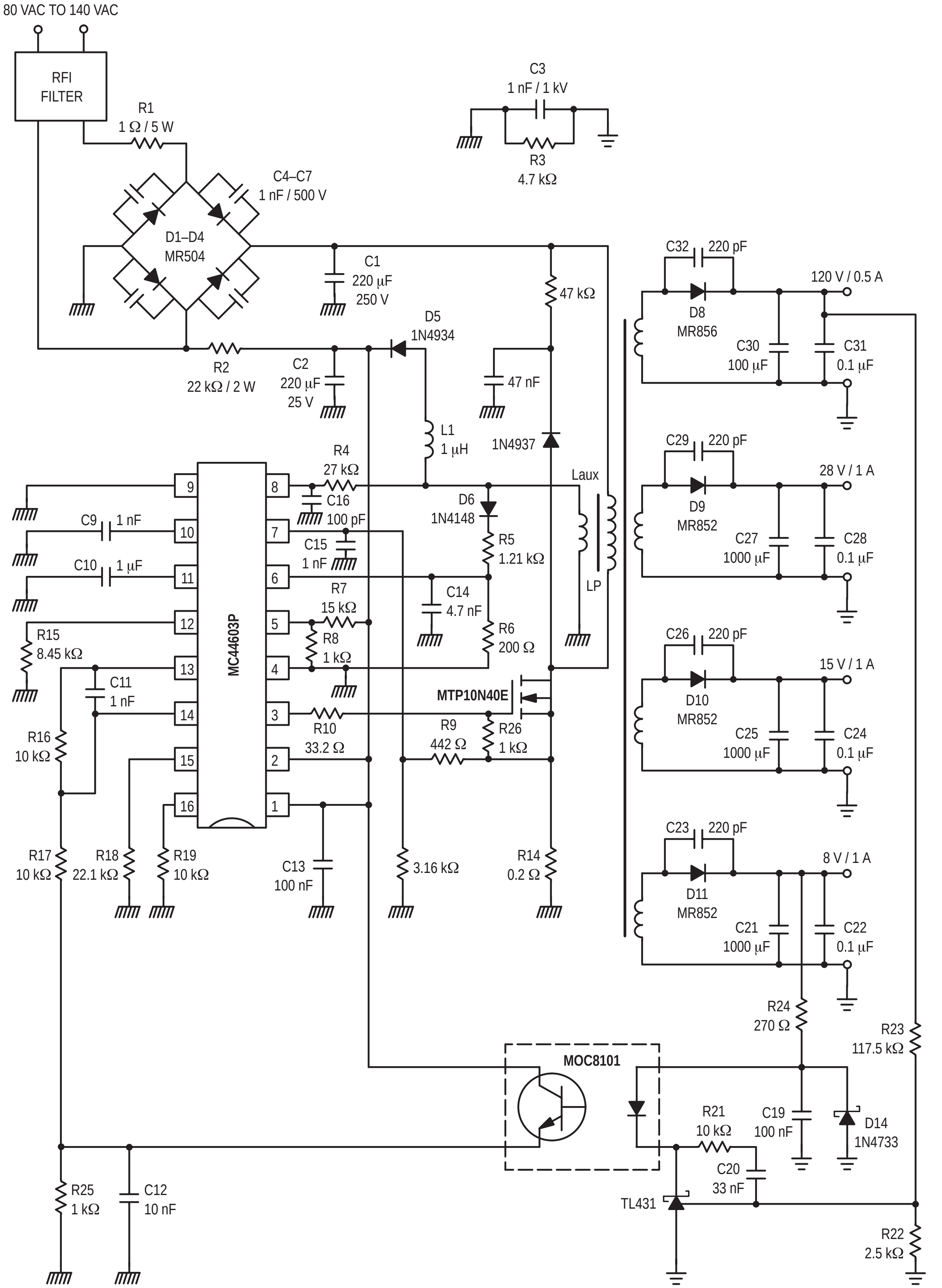


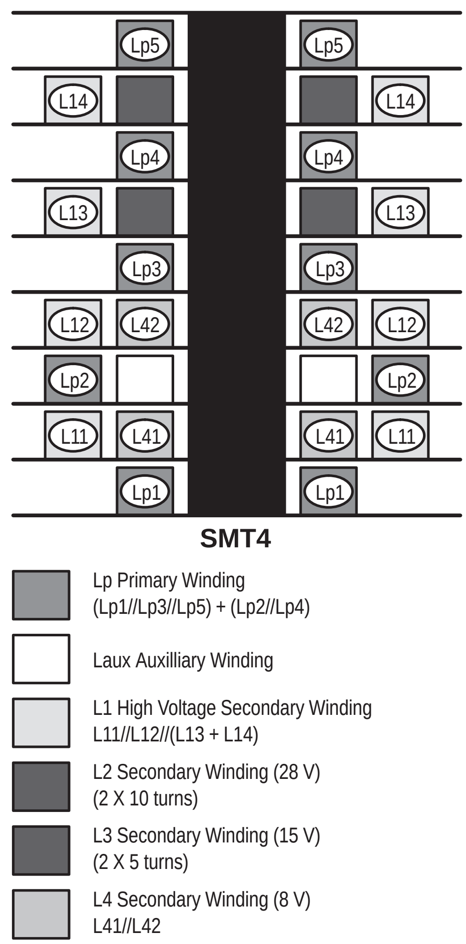
Figure 4. 110 W Output Off–Line Flyback Converter with MOSFET Switch, 80 V ↔ 140 V Mains Voltage

Table 1. 110 W Fly–Back Converter, 80 Vrms–140 Vrms Mains Range, MC44603 and MTP10N40E

Test		Conditions	Results
Line Regulation		Vin = 90 Vac to 140 Vac Fmains = 50 HZ Fmains = 50 HZ	
	120 V	Iout = 0.5A	Δ = 0 V
	28 V	Iout = 1A	Δ = 0 V
	15 V	Iout = 1A	Δ = 0 V
	8 V	Iout = 1A	Δ = 0 V
Load Regulation		Vin = 110 Vac	
	120 V	Iout = 0.3A to 0.5A	Δ = 0.05 V
Cross Regulation		Vin = 110 Vac	
		Iout (120 V) = 0.5A	
		Iout (28 V) = 0A to 1A	
	120 V	Iout (15 V) = 1A	Δ = 0 V
		Iout (8 V) = 1A	
Efficiency		Vin = 110 Vac, Po = 110 W	84.5%
Standby Mode			
P input		Vin = 110 Vac, Pout = 0 W	1.2 W
Switch. freq.			20 KHz fully stable
Output short circuit			Safe on all outputs
Start–up		Pin 110 W	Vac = 80 V

MOSFET application: information about the transformer

	110 W
Lp (turns)	30
Laux (turns)	5
L1 (turns)	40
L2 (turns)	10
L3 (turns)	5
L4 (turns)	3
Al (nH/turns^2)	274
Core	E–4215A
Material	B2
Former	specific Thomson design
Wire size (mm^2)	0.315 all windings



Flyback transformer construction

For cost reduction and simplicity, all windings have the same size. For optimal Lp/Laux coupling, Laux is wound on the second section of Lp.

Former

The normalized primary/secondary isolation is obtained using the multi–slotted former depicted on the figure. This former uses designs patented by LCC Thomson.

IV — APPLICATION 2: 220 V INPUT VOLTAGE

IV–1 — Choice of the transformer:

One way to use the above design equations, consists of drawing up a table showing how the main SMPS parameters vary with the value of the turn ratio.

To calculate these values, it is necessary to know the input power level. This value is taken equal to (135 W) in our application (135 W corresponds to an efficiency equal to about 80%. The application results will show that this assumption ensures a desirable margin with the nominal input voltage).

On the other hand, the parameters calculation shows that (L x fosc)max is the (L x fosc) value that results in the lowest (Ipk)max and (pon)max ones (refer to Ipk or pon expressions). This (L x fosc) value is the maximum one that guarantees a fixed frequency working for any working point (refer to section II–1). The SMPS parameters given in the following table are calculated using this threshold value.

Choice criteria and definition of the transformer:

As shown by the following table, the higher the turn ratio (N) is, the lower the peak current is. Now, the (ni)max is proportional to N and the voltage the transistor must face, increases when N rises. That is why an optimal N value must be chosen.

MOSFET case:

To perform a low cost SMPS, it is required to use a MOSFET 600 V. It is necessary to have a safety voltage margin, to avoid the need to incorporate a lossy and costly clamping network that would cut the voltage spikes due to the leakage inductor at the power switch turning off (refer to Figure 1 in section III–1).

Practically, about 550 V is acceptable. Consequently, (N = 1.2) seems to be a maximum value.

Now, in order to obtain a well coupled transformer with a low leakage inductor value, it is desirable to use a ferrite with a low air–gap.

So, in order to be able to use a ferrite (ni = 140, A_L = 274 nH/turns^2), (N = 1) seems to be a preferable value.

Consequently,

$$Lp = A_L \times (N \times 40)^2 \longrightarrow Lp \approx 438\mu H$$

and the optimal working frequency is:

$$L \times fosc \leq 24.3 \longrightarrow fosc \leq 55 \text{ kHz}$$

So, the following values can be chosen:

L = 438 µH
fosc = 50 kHz

and then Ipk = 3.5A

BIPOLAR transistor case:

As the gain of a Bipolar transistor decreases when the collector current level rises, the SMPS peak current must be as low as possible. That is why N must be chosen as high as possible. Now, if classical BIPOLAR transistors are able to face 1000 V or 1200 V, their V_{CEO} is generally low. The transistor used in the application, the MJE18206, has a V_{CES} equal to 1200 V and a V_{CEO} equal to 600 V. Since there are damped oscillations (converging to Vin) during the dead– time (refer to Figure 1), the transistor may be turned on while its V_{CE} voltage is higher than Vin (the maximum Vin value being nearly equal to 400 V). That is why, even if a resistor is connected between the base and the emitter of the transistor (refer to section IV–2), the $(V_T)_{max}$ (that is, $(Vin+NVo)_{max}$) must be chosen lower than 600 V, to ensure system reliability.

In addition to this, a second choice criterion is (ni)max, since transformer saturation must be avoided.

(N = 1.6) seems to be a good choice that enables the use of a ferrite (A_L = 250nH/turns^2; ni = 180))

Consequently,

$$L = A_L \times (N \times 40)^2 \longrightarrow L \approx 1mH$$

So, the optimal working frequency is:

$$L \times fosc \leq 43.7 \longrightarrow fosc \leq 43 \text{ kHz}$$

Finally, the following value can be taken:

L = 1mH
fosc = 43 kHz
(Ipk)max = 2.5A

N	(L.fosc)max	(Ipk)max (A)	(V_T)max (V)	(V_D)max (V)	MOSFET on losses/Rdson (W/Ω)	BIPOLAR on losses/V_{CE} (W/A)	(ni)max
0.75	16.2	4.1	490	650	1.5	0.54	122
1.00	24.3	3.3	520	520	1.2	0.54	133
1.20	30.9	3.0	540	450	1.1	0.54	144
1.40	37.4	2.7	570	400	1.0	0.54	150
1.60	43.7	2.5	590	370	0.9	0.54	159
1.80	49.7	2.3	620	340	0.8	0.54	168
2.00	55.5	2.2	640	320	0.8	0.54	176

NOTE: N: turn ratio (refer to II–5)
(V_T)max: maximum voltage the power switch must face
(V_D)max: maximum voltage the 120 V output diode must face

IV–2 — MC44603 pins use: (refer to the application schematics)

1 — Vcc (PIN 1):

The pin Vcc must be connected to a transformer auxilliary winding. This extra winding turns number can be chosen equal to 5, in order to obtain a Vcc nearly equal to 15 V.

2 — Vc and OUTPUT (pins 2 and 3):

Vc is the output high state of the circuit. This pin offers the possibility of setting the output source current at a different level than the sink current.

— MOSFET case:

A resistor of 10 Ω must be connected between the output and the MOSFET gate to make the switchings smoother. A resistor of about 1 kΩ can be connected between the gate and the ground (or the current sense external resistor) to avoid any inadvertent MOSFET switching on due to noise.

— BIPOLAR transistor case:

For the on–time, a bipolar transistor requires a base current labelled I_{B1}, that must be higher than:

(Ic)max / ßmin

where

(Ic)max is the maximum collector current (that is Ipk_{max} if the current sense resistor is well designed — refer to section IV–5), and

ßmin is the minimum guaranteed transistor gain for (Ic = (Ic)max)

Now, with the MJE18206: (ßmin ≈ 7) for (Ic)max = 2.5A

So, (I_{B1} = 400mA) is a good value that ensures a safety margin.

On the other hand, the turn off base current peak must be nearly equal to (2 x I_{B1}). The couple (Dz,Cz) is used to build a voltage source Vz (during the on–time), able to produce I_{B2}.

So, I_{B1} = (Vcc – Vz – Vbe) / (r1 + r2)

I_{B2} = (Vz + Vbe) / r2

Consequently, using Vz = 3.3 V
Cz = 1 µF
r1 = 22 Ω
r2 = 4.7 Ω

As (Vcc ≈ 15 V), the obtained base currents are:

I_{B1} ≈ 410 mA

I_{B2} ≈ 850 mA

These base currents enable a correct transistor drive.

3 — Foldback (pin 5):

A portion of Vcc must be applied to this pin thanks to a resistor divider. This voltage value must be slightly higher than 1 V in normal working so that this value drops below this threshold value as soon as an overload occurs.

4 — Overvoltage protection (pin 6):

This pin can remain free and then, the Vcc threshold level is fixed equal to nearly 17 V.

On the other hand, to make detection quicker and more accurate, an external resistor divider can be used with a diode and an integration capacitor. In the proposed application, this resistor divider is not directly connected to the Vcc because Vcc has a high time constant (refer to the application schematics).

5 — Current sense (pin 7):

The current sense resistor must be designed in order to limit the current down to the maximum peak calculated in section II in order to limit the power the converter is able to draw from the mains (in a fixed frequency mode, Pin = 1/ 2 x L x Ipk^2 x fosc).

— MOSFET case: (Ipk)max = 3.5A

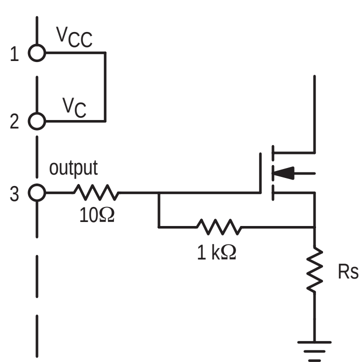
Now, the (Vcs) clamp level is nearly 1 V (refer to the data sheet).

So, (Rs) the current sense resistor, must be equal to (1 V / 3.5A), that is nearly: 0.28Ω (2 x 0.56 Ω in parallel).

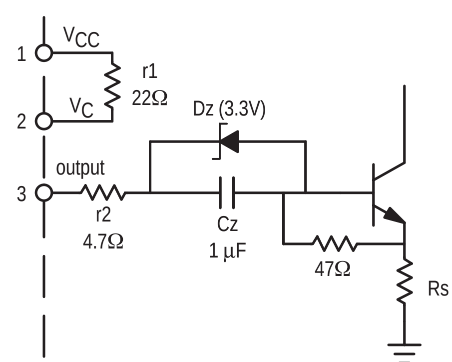
— BIPOLAR case: (Ipk)max = 2.5A

So, (Rs) must be equal to (1 V / 2.5A), that is: 0.4 Ω (3 x 1.2 Ω in parallel).

Finally, as the fixed frequency mode is ensured for any working point, the peak current limitation results in an accurate input power limitation (135 W in this application).



MOSFET drive



BIPOLAR transistor drive

Figure 5.

6 — Oscillator (pin 10 & 16):

The oscillator frequency is determined by the couple (C_T, Rref) (refer to the data sheet).

As capacitors have discrete values, the choice of Rref allows you to fix precisely the oscillator frequency (however, Rref also fixes the internal current source (Iref), which must be lower than 500 µA and higher than 100 µA).

MOSFET case:
fosc = 50 kHz ⟶ Rref = 10 kΩ ⟶ C_T = 820pF

BIPOLAR case:
fosc = 43 kHz ⟶ Rref = 10 kΩ ⟶ C_T = 1nF

7 — Stand–by mode (pins 12 and 15):

In the MC44603, it is possible to reduce the working frequency when little power is being drawn from the mains (stand–by mode).

This stand–by frequency is fixed by connecting a resistor R_{Fstby} to pin 15, while the power level at which the stand–by mode must be applied is determined by connecting another resistor R_{Pstby} to pin 12 (this power level is labelled P_{thL} in the data sheet).

In the data sheet, the equations needed to calculate R_{Fstby} and R_{Pstby} are indicated.

Using them, to obtain a power level equal to 15 W and a stand–by frequency equal to 20 kHz, the calculated R_{Fstby} and R_{Pstby} values are:

MOSFET case:	RPstby = 10 kΩ	RFstby = 27 kΩ
BIPOLAR case:	RPstby = 10 kΩ	RFstby = 22 kΩ

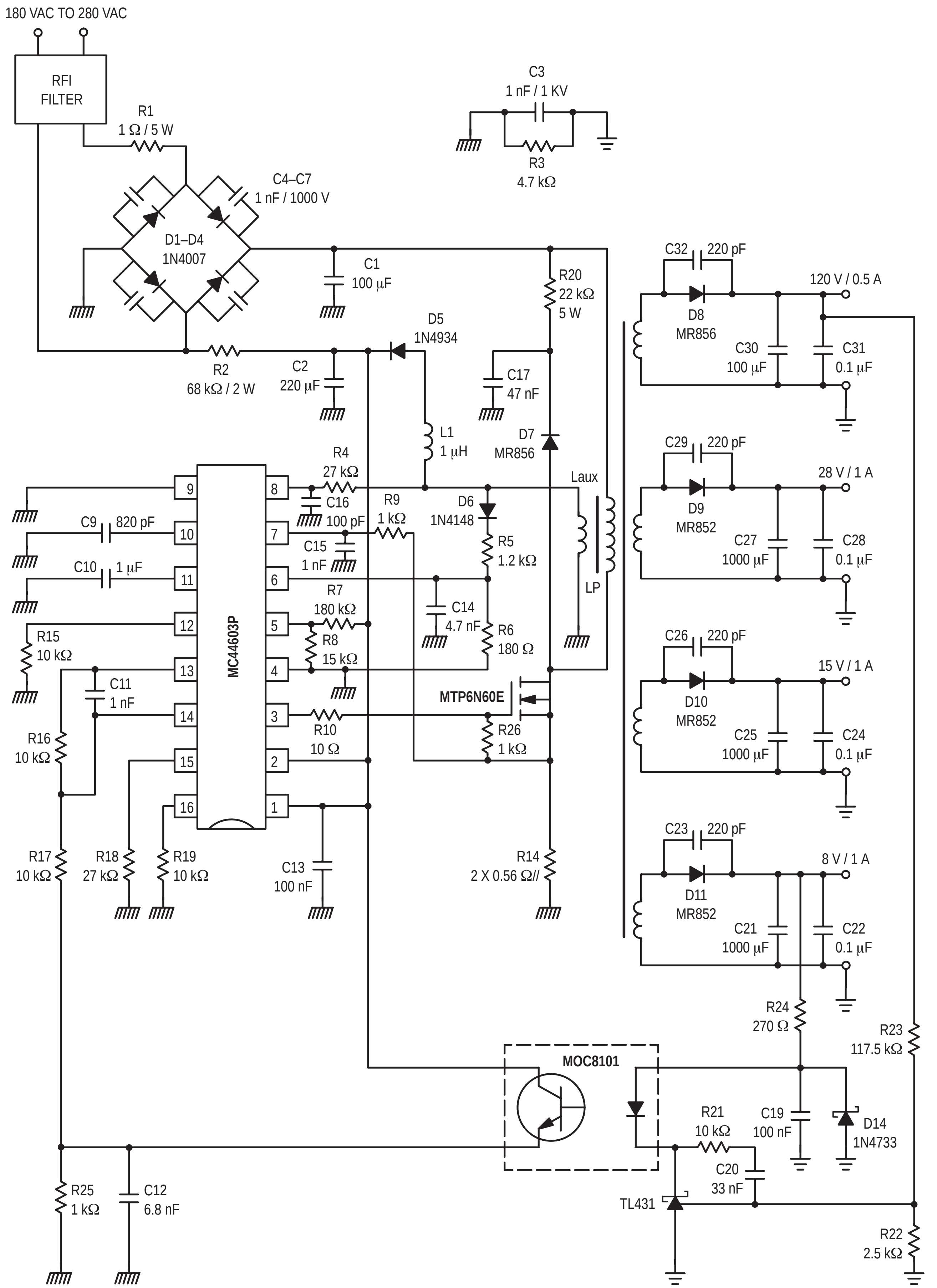


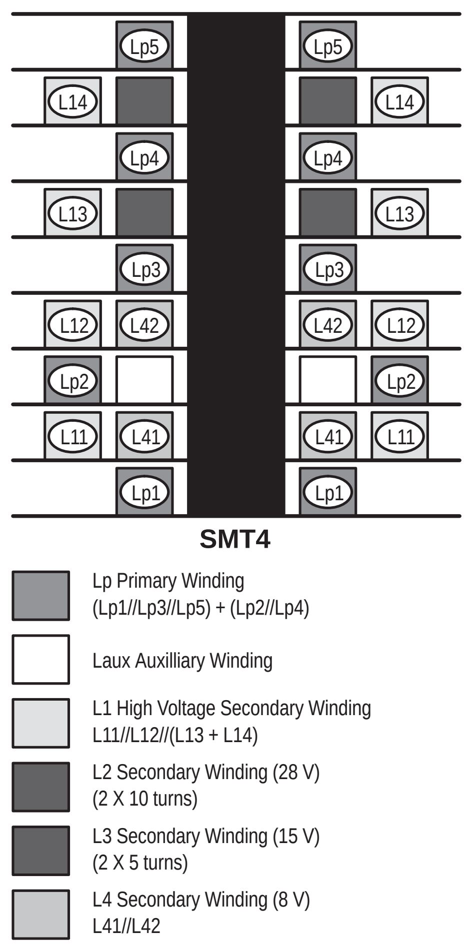
Figure 6. 110 W Output Off–Line Flyback Converter with MOSFET Switch. 180 V–280 V MAINS RANGE

Table 2. 110 W Fly–Back Converter, 180 V–280 V Mains Range, MC44603 and MTP6N60E

Test		Conditions	Results
Line Regulation		Vin = 180 Vac to 280 Vac Fmains = 50 HZ	
	120 V	Iout = 0.5A	Δ = 0 V
	28 V	Iout = 1A	Δ = 0 V
	15 V	Iout = 1A	Δ = 0 V
	8 V	Iout = 1A	Δ = 0 V
Load Regulation		Vin = 220 Vac	
	155 V	Iout = 0.3A to 0.5A	Δ = 0.05 V
Cross Regulation		Vin = 220 Vac Iout (120 V) = 0.5A Iout (28 V) = 0A to 1A	
	120 V	Iout (15 V) = 1A Iout (8 V) = 1A	Δ = 0 V
Efficiency		Vin = 220 Vac, Po = 110 W	84%
Standby Mode P input Switch. freq.		Vin = 220 Vac, Pout = 0 W	3 W 20 KHz fully stable
Output short circuit			Safe on all outputs
Start–up		Pin 110 W	Vac = 160 V

MOSFET application: information about the transformer

	110 W
Lp (turns)	40
Laux (turns)	5
L1 (turns)	40
L2 (turns)	10
L3 (turns)	5
L4 (turns)	3
Al (nH/turns^2)	274
Core	E–4215A
Material	B2
Former	specific Thomson design
Wire size (mm^2)	0.315 all windings



Flyback transformer construction

For cost reduction and simplicity, all windings have the same size. For optimal Lp/Laux coupling, Laux is wound on the second section of Lp.

Former

The normalized primary/secondary isolation is obtained using the multi–slotted former depicted on the figure. This former uses designs patented by LCC Thomson.

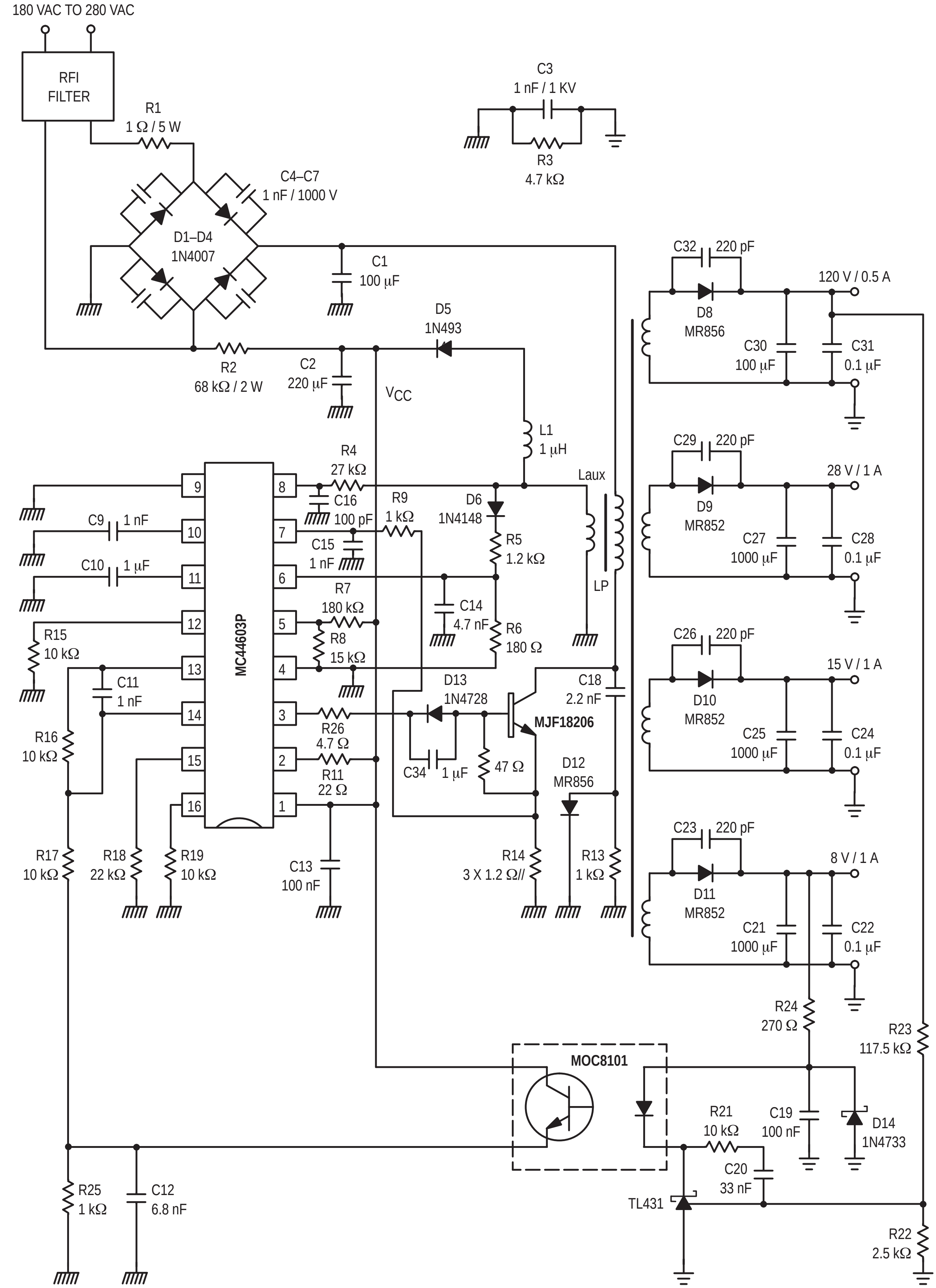


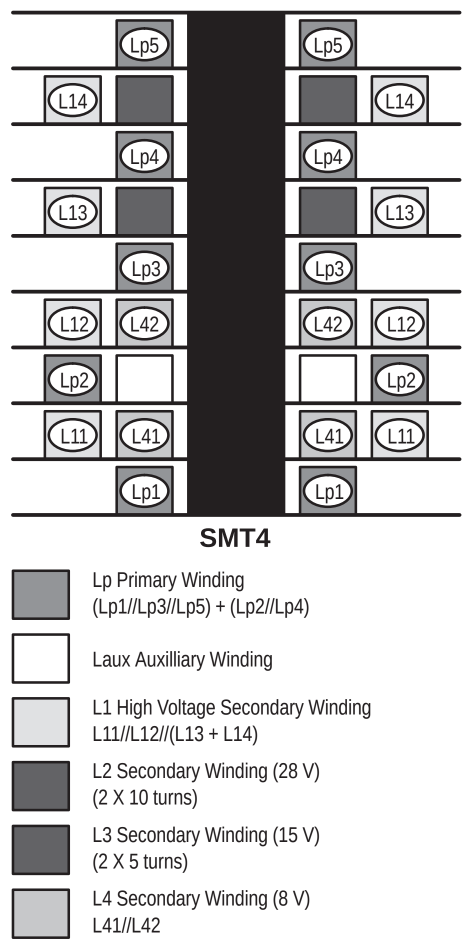
Figure 7. 110 W Output Off–Line Flyback Converter with Bipolar Switch. 180 V–280 V MAINS RANGE

Table 3. 110 W Fly–Back Converter, 180 V–280 V Mains Range, MC44603 and MJF18206

Test	Conditions	Results
Line Regulation	Vin = 180 Vac to 280 Vac Fmains = 50 HZ	
120 V	Iout = 0.5A	Δ = 0 V
28 V	Iout = 1A	Δ = 0 V
15 V	Iout = 1A	Δ = 0 V
8 V	Iout = 1A	Δ = 0 V
Load Regulation	Vin = 220 Vac	
120 V	Iout = 0.2A to 0.5A	Δ = 0.05 V
Cross Regulation	Vin = 220 Vac Iout (120 V) = 0.5A Iout (28 V) = 0A to 1A	
120 V	Iout (15 V) = 1A Iout (8 V) = 1A	Δ = 0 V
Efficiency	Vin = 220 Vac, Po = 110 W	85%
Standby Mode P input Switch. freq.	Vin = 220 Vac, Pout = 0 W	3W 20 KHz fully stable
Output short circuit		Safe on all outputs
Start–up	Pin 110 W	Vac = 160 V

BIPOLAR application: information about the transformer

	110 W
Lp (turns)	64
Laux (turns)	5
L1 (turns)	40
L2 (turns)	10
L3 (turns)	5
L4 (turns)	3
Al (nH/turns^2)	250
Core	E–4215A
Material	B2
Former	specific Thomson design
Wire size (mm^2)	0.315 all windings



Flyback transformer construction

For cost reduction and simplicity, all windings have the same size. For optimal Lp/Laux coupling, Laux is wound on the second section of Lp.

Former

The normalized primary/secondary isolation is obtained using the multi–slotted former depicted on the figure. This former uses designs patented by LCC Thomson.

V — CONCLUSION

These applications show a significant advantage of the fixed frequency mode: it enables us to precisely limit the maximum power that may be drawn by the converter from the mains (135 W in our case).

Note that the stand by losses are lower in the 110 V application because in this case, only a low loss, costly clamping network is used to protect the MOSFET (no snubber). Indeed, the snubber and clamping arrangements dissipate some energy (that is not insignificant) at each switching. That is why the reduction of the switching frequency is a very effective means to decrease the stand–by losses (the snubber and clamping arrangement cannot be removed in most cases).

This application note does not pay much attention to the MC44603's protection features. Two features are especially noteworthy:

— the foldback that protects the converter when there is an overload

— the effective demagnetization section that ensures a discontinuous mode

Notes:

— the MOSFET on time losses are high in the 110 V application. The use of a MOSFET having a lower Rdson (or two MOSFET in parallel) would improve the efficiency

— because of these losses, the input range of the proposed solution is actually: 90 V–140 V

In order to minimize the length of this application note, it does not consider a universal mains range application. Such a SMPS could be designed using the methods described here.

APPENDIX

OREGA TRANSFORMERS

Type	A_L	(ni) @100°C	Ferrite	Wire ∅ (mm)	Nmax
SMT1	448	40	5H20	0.25	56
	260	80	5H20	0.25	56
	240	85	5H20	0.224	68
	220	90	5H20	0.224	68
	180	125	5H20	0.224	68
SMT3	350	80	B1	0.315	75
	250	110	B1	0.315	75
	250	130	B3	0.315	75
	190	160	B1	0.315	75
	178	180	B1	0.315	75
SMT4	336	110	B1	0.28	68
	320	135	B3	0.315	52
	274	140	B1	0.40	36
	250	180	B3	0.40	36
	238	200	B3	0.40	36
	215	190	B1	0.40	36
	192	210	B1	0.40	36
	192	245	B3	0.40	36
SMT47	560	100	B3	0.315	76
	428	140	B1	0.40	46
	428	150	B3	0.50	26
	372	190	B3	7 X 0.2	20
	315	220	B3	7 X 0.2	20
	262	270	B1	7 X 0.2	20
	234	310	B3	7 X 0.2	20

The last column indicates the maximum number of turns per slot (refer to page 14) that will fall within the insulation norms, when using wires whose size is indicated in the "wire" column.

NOTES

NOTES

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