

## AN4001/D

# MC44603 in a 110W output SMPS application (80–140Vrms and 180–280Vrms 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 110W output power supply in two distinct cases: 110Vrms mains and 220Vrms 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: 110W

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

Application 1: 110V mains 80Vrms  $\leftrightarrow$  140Vrms

Application 2: European mains 180Vrms  $\leftrightarrow$  280Vrms

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 1V, 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 9V 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 9V should be well coupled; also, 40 turns (for the 120V) 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 120V 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:

$$T_{on} + T_{off} \leq T_{osc} \quad \text{ineq 1}$$

where:  $T_{on}$  is the on-time,  $T_{off}$  is the off-time  
 $T_{osc}$  is the MC44603 oscillator period

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

So, 
$$I_{pk} = \frac{V_{in}}{L} \times T_{on} \quad \text{eqn 1}$$

and, 
$$I_{pk} = \frac{NV_o}{L} \times T_{off} \quad \text{eqn 2}$$

where:  $I_{pk}$  is the primary inductor peak current  
 $V_{in}$  is the rectified a.c. line voltage (fly-back input voltage)  
 $L$  is the primary inductor value  
 $NV_o$  is the output voltage considered in the primary side  
( $N$ : turn ratio)

So, 
$$T_{on} + T_{off} = L \times I_{pk} \times \frac{V_{in} + NV_o}{V_{in} \times NV_o} \quad \text{eqn 3}$$

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

$$E = \frac{1}{2} \times L \times I_{pk}^2 \quad \text{eqn 4}$$

So, Pin (input power) is: 
$$P_{in} = \frac{1}{2} \times L \times I_{pk}^2 \times f_{osc}$$
 eqn 5

where  $f_{osc}$  is the MC44603 oscillator frequency

Using equations 3 and 5 and inequality 1,

**condition required to work in a fixed frequency mode:**

$$L \times f_{osc} \leq \frac{1}{2 \times P_{in}} \times \left( \frac{V_{in} \times N V_o}{V_{in} + N V_o} \right)^2$$
 ineq 2

## II - 2 - Peak inductor current expression

Pin is given by the following expression:

$$P_{in} = V_{in} \times I_{in}$$
 eqn 6

where  $I_{in}$  is the input current

and 
$$I_{in} = \frac{1}{2} \times I_{pk} \times d$$
 eqn 7

where  $d$  is the duty cycle

Now, 
$$T_{on} = d \times T_{osc}$$
 eqn 8

Using equations 1, 7 and 8:

$$I_{in} = \frac{1}{2} \times L \times I_{pk}^2 \times \frac{f_{osc}}{V_{in}}$$
 eqn 9

Using equations 6 and 9:

$$I_{pk} = \sqrt{\frac{2 \times P_{in}}{L \times f_{osc}}}$$
 eqn 10

Thus, Peak inductor current:

$$(I_{pk})_{max} = \sqrt{\frac{2 \times (P_{in})_{max}}{L \times f_{osc}}}$$
 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:

$$P_{on} = \left( \frac{1}{T_{osc}} \right) \times \int_0^{T_{on}} R_{dson} \times I_T^2 \times dt \quad \text{eqn 12}$$

where  $I_T$  is the MOSFET current.

So, as: 
$$I_T = V_{in} \times \frac{t}{L} \quad \text{eqn 13}$$

$$P_{on} = \frac{1}{3} \times R_{dson} \times V_{in}^2 \times \frac{T_{on}^3}{(T_{osc} \times L^2)} \quad \text{eqn 14}$$

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

$$P_{on} = \frac{2 \times \sqrt{2}}{3} \times R_{dson} \times \frac{P_{in}}{V_{in}} \times \sqrt{\frac{P_{in}}{L \times f_{osc}}} \quad \text{eqn 15}$$

Consequently,

#### **maximum MOSFET on-time losses**

$$(P_{on})_{max} = \frac{2 \times \sqrt{2}}{3} \times R_{dson} \times \frac{(P_{in})_{max}}{(V_{in})_{min}} \times \sqrt{\frac{(P_{in})_{max}}{L \times f_{osc}}} \quad \text{eqn 16}$$

Using the  $(I_{pk})_{max}$  value, the following equation could also be written:

$$(P_{on})_{max} = \frac{1}{3} \times R_{dson} \times (I_{pk})_{max}^2 \times d_{max}$$

where  $d_{max}$  is the maximum duty cycle.

### II - 3 - 2 - BIPOLAR TRANSISTOR

$$P_{on} = \frac{1}{T_{osc}} \times \int_0^{T_{on}} V_{CE} \times I_T \times dt \quad \text{eqn 17}$$

So:  $P_{on} = V_{CE} \times \langle I_T \rangle \quad \text{eqn 18}$

Thus:  $P_{on} = V_{CE} \times I_{in} \quad \text{eqn 19}$

where  $I_{in}$  is the input current

So:  $P_{on} = V_{CE} \times \frac{P_{in}}{V_{in}} \quad \text{eqn 20}$

Consequently,

**maximum BIPOLAR transistor on-time losses**

$$(P_{on})_{max} = V_{CE} \times \frac{(P_{in})_{max}}{(V_{in})_{min}} \quad \text{eqn 21}$$

### II - 4 - Maximum duty cycle

The duty cycle,  $d$ , is equal to:  $d = \frac{T_{on}}{T_{osc}} \quad \text{eqn 22}$

Now,  $T_{on} = L \times \frac{I_{pk}}{V_{in}} \quad \text{eqn 23}$

so,  $d = L \times f_{osc} \times \frac{I_{pk}}{V_{in}} \quad \text{eqn 24}$

Thus, using equation 10:  $d = \sqrt{2 \times P_{in} \times L \times \frac{f_{osc}}{V_{in}^2}} \quad \text{eqn 25}$

So, **maximum duty cycle:**

$$d_{\max} = \sqrt{2 \times (P_{\text{in}})_{\max} \times L \times \frac{f_{\text{osc}}}{(V_{\text{in}})_{\min}^2}} \quad \text{eqn 26}$$

## 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:**

$$(V_T)_{\max} = (\sqrt{2} \times (V_{\text{in}})_{\max}) + (N \times 120) \text{ (V)} \quad \text{eqn 27}$$

where:  $(V_{\text{in}})_{\max}$  is the maximum rms a.c. line voltage (140V or 280V according to the line)  
120 is the maximum output voltage  
N is the transformer turn ratio between the 120V output winding and the primary inductor

**$(V_D)_{\max}$ : the maximum voltage the 120V output diode must face:**

$$(V_D)_{\max} = \left( \sqrt{2} \times \frac{(V_{\text{in}})_{\max}}{N} \right) + 120 \text{ (V)} \quad \text{eqn 28}$$

**$(ni)$ : 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 120V winding

So, 
$$ni = N \times n_{120V} \times I_{\text{pk}} \quad \text{eqn 30}$$

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  $(f_{\text{osc}} \times L)$ .

### III - 2 - MC44603 pins use: (refer to the application schematics)

#### 1 - Vcc (PIN 1):

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

#### 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  $33.2\Omega$  must be connected between the output and the MOSFET gate to make the switchings smoother. A resistor of about  $1k\Omega$  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).

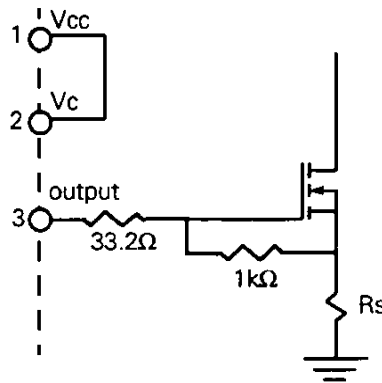


Figure 2.

#### 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 1V 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 17V.

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,  $P_{in} = 1/2 \times L \times I_{pk}^2 \times f_{osc}$ .

Now,  $(I_{pk})_{max} = 5.4A$

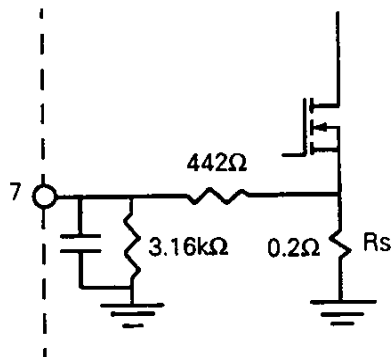


In addition, the (Vcs) clamp level is nearly 1V (refer to the data sheet)

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

$$(1V / 5.4A), \text{ that is nearly: } 0.18\Omega.$$

This value can be obtained using a 1W, 0.2 $\Omega$  resistor and a resistor divider (442 $\Omega$ , 3.16k $\Omega$ ) (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 (135W in this application note).

#### 6 - Oscillator (pins 10 and 16):

The oscillator frequency is determined by the couple ( $C_T$ ,  $R_{ref}$ ) (refer to the data sheet).

Even if capacitors have discrete values, the choice of  $R_{ref}$  allows you to fix precisely the oscillator frequency (however,  $R_{ref}$  also fixes the internal current source ( $I_{ref}$ ), which must be lower than 500 $\mu$ A and higher than 100 $\mu$ A).

$$\begin{aligned} f_{osc} = 40\text{kHz} & \longrightarrow R_{ref} = 10\text{k}\Omega \\ & \longrightarrow C_T = 1\text{nF} \end{aligned}$$

#### 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 10W and a stand-by frequency equal to 20kHz, the calculated  $R_{Fstby}$  and  $R_{Pstby}$  values are:

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

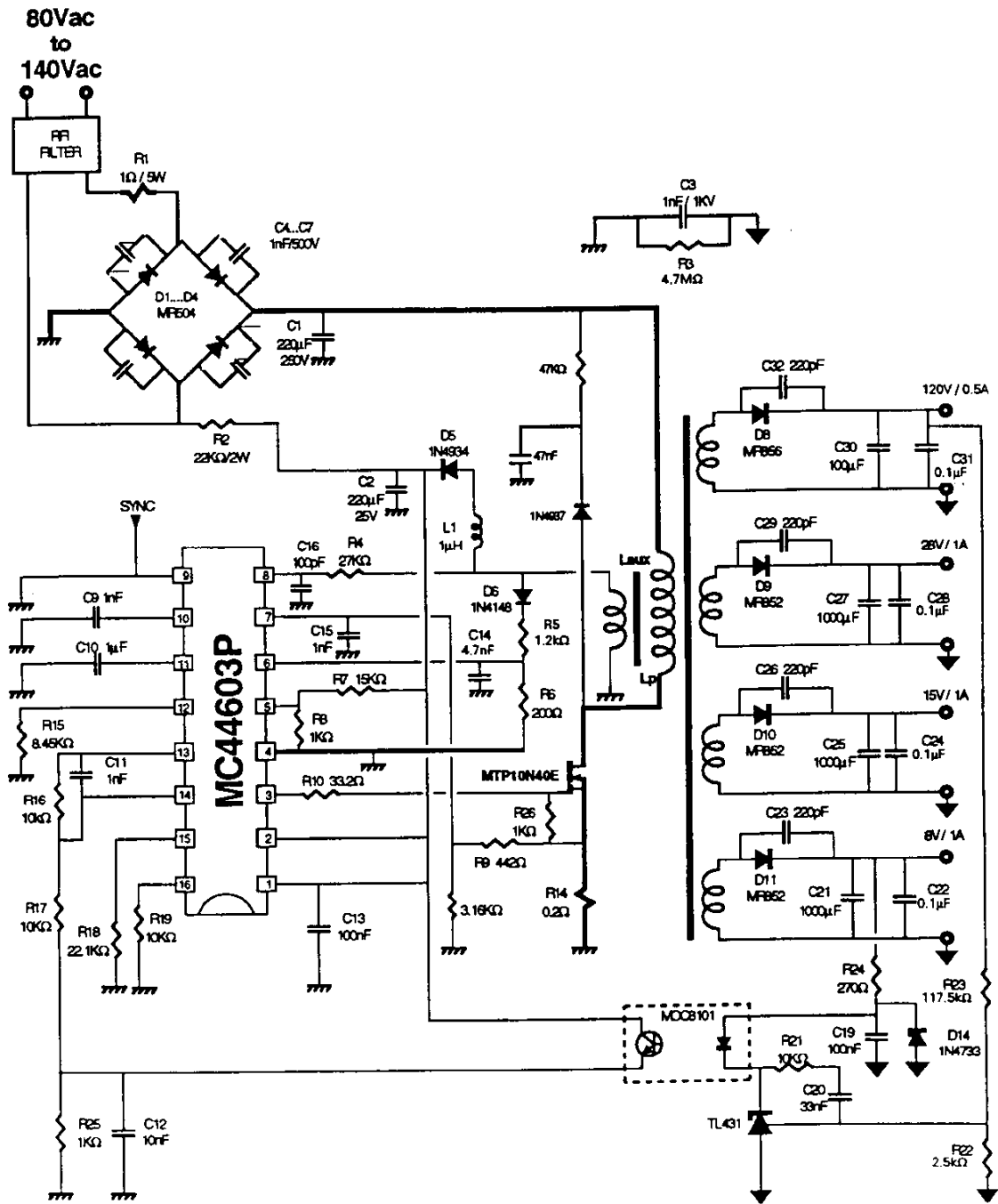


Figure 4. 110W output Off-Line Flyback Converter with MOSFET Switch, 80V ↔ 140V mains voltage

**Table 1.** 110W fly-back converter, 80Vrms – 140Vrms mains range, MC44603 and MTP10N40E

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

## MOSFET application: information about the transformer

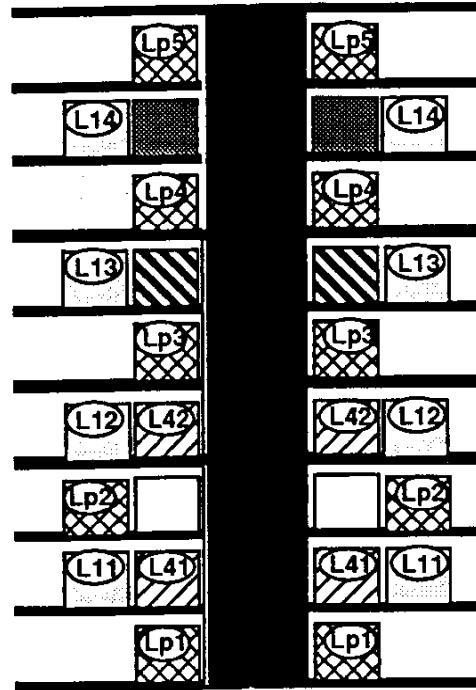
	110W
Lp (turns)	30
Laux (turns)	5
L1 (turns)	40
L2 (turns)	10
L3 (turns)	5
L4 (turns)	3
Al (nH/turns <sup>2</sup> )	274
Core	E-4215A
Material	B2
Former	specific Thomson design
Wire size (mm <sup>2</sup> )	0.315 all windings

### 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.



SMT4

-  Lp Primary winding  
(Lp1//Lp3//Lp5)+(Lp2//Lp4)
-  Laux Auxilliary winding
-  L1 High voltage secondary winding  
L11//L12//(L13+L14)
-  L2 Secondary winding (28V)  
(2 x 10 turns)
-  L3 Secondary winding (15V)  
(2 x 5 turns)
-  L4 Secondary winding (8V)  
L41//L42

## IV- APPLICATION 2: 220V 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 (135W) in our application (135W 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  $(L \times f_{osc})_{max}$  is the  $(L \times f_{osc})$  value that results in the lowest  $(I_{pk})_{max}$  and  $(V_{T})_{max}$  ones (refer to  $I_{pk}$  or  $V_{T}$  expressions). This  $(L \times f_{osc})$  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.

N	$(L.f_{osc})_{max}$	$(I_{pk})_{max}$ (A)	$(V_T)_{max}$ (V)	$(V_D)_{max}$ (V)	MOSFET on losses/ $R_{dson}$ (W/ $\Omega$ )	BiPOLAR on losses/ $V_{CE}$ (W/A)	$(n_i)_{max}$
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 120V output diode must face

### Choice criteria and definition of the transformer:

As shown by the above table, the higher the turn ratio (N) is, the lower the peak current is. Now, the  $(n_i)_{max}$  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 600V. 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 550V 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 ( $n_i = 140$ ,  $A_L = 274 \text{ nH/turns}^2$ ), (N = 1) seems to be a preferable value.

Consequently,  $L_p = A_L \times (N \times 40)^2 \longrightarrow L_p \approx 438\mu\text{H}$

and the optimal working frequency is:

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

So, the following values can be chosen:

$$L = 438\mu\text{H}$$

$$f_{osc} = 50\text{kHz}$$

and then

$$I_{pk} = 3.5\text{A}$$

#### *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 1000V or 1200V, their  $V_{CEO}$  is generally low. The transistor used in the application, the MJE18206, has a  $V_{CES}$  equal to 1200V and a  $V_{CEO}$  equal to 600V. Since there are damped oscillations (converging to  $V_{in}$ ) during the dead-time (refer to figure 1), the transistor may be turned on while its  $V_{CE}$  voltage is higher than  $V_{in}$  (the maximum  $V_{in}$  value being nearly equal to 400V). 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,  $(V_{in} + NV_o)_{max}$ ) must be chosen lower than 600V, to ensure system reliability.

In addition to this, a second choice criterion is  $(n_i)_{max}$ , since transformer saturation must be avoided.

(N=1.6) seems to be a good choice that enables the use of a ferrite ( $A_L = 250\text{nH/turns}^2$ ;  $n_i=180$ )

Consequently,  $L = A_L \times (N \times 40)^2 \longrightarrow L \approx 1\text{mH}$

So, the optimal working frequency is:

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

Finally, the following value can be taken:

$$L = 1\text{mH}$$

$$f_{osc} = 43\text{kHz}$$

$$(I_{pk})_{max} = 2.5\text{A}$$

**IV - 2 - MC44603 pins use: (refer to the application schematics)**

**1 - Vcc (PIN 1):**

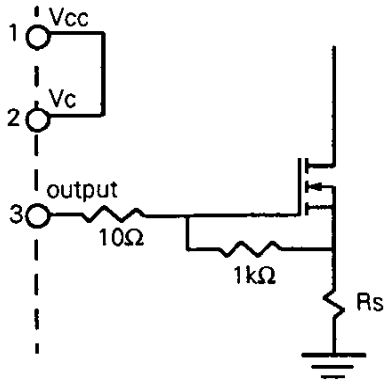
The pin Vcc must be connected to a transformer auxilliary winding. This extra winding turns number can be chosen equal to 5, in order to obtain a Vcc nearly equal to 15V.

**2 - Vc and OUTPUT (pins 2 and 3):**

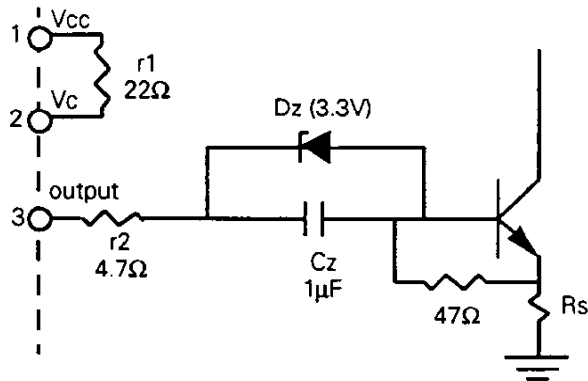
Vc is the output high state of the circuit. This pin offers the possibility of setting the output source current at a different level than the sink current.

**- MOSFET case:**

A resistor of 10Ω must be connected between the output and the MOSFET gate to make the switchings smoother. A resistor of about 1kΩ 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.



MOSFET drive



BIPOLAR transistor drive

**Figure 5.**

**- BIPOLAR transistor case:**

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

$$(I_c)_{max} / \beta_{min}$$

where

$(I_c)_{max}$  is the maximum collector current (that is  $I_{pk_{max}}$  if the current sense resistor is well designed - refer to section IV - 5), and  $\beta_{min}$  is the minimum guaranteed transistor gain for  $(I_c = (I_c)_{max})$

Now, with the MJE18206: ( $\beta_{min} = 7$ ) for  $(I_c)_{max} = 2.5A$

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  $(2 \times I_{B1})$ . The couple  $(Dz, Cz)$  is used to build a voltage source  $Vz$  (during the on-time), able to produce  $I_{B2}$ .

So, 
$$I_{B1} = (V_{cc} - Vz - V_{be}) / (r1 + r2)$$

$$I_{B2} = (Vz + V_{be}) / r2$$

Consequently, using 
$$Vz = 3.3V$$

$$Cz = 1\mu F$$

$$r1 = 22\Omega$$

$$r2 = 4.7\Omega$$

As  $(V_{cc} \approx 15V)$ , the obtained base currents are:

$$I_{B1} \approx 410 \text{ mA}$$

$$I_{B2} \approx 850 \text{ mA}$$

These base currents enable a correct transistor drive.

### 3- Foldback (pin 5):

A portion of  $V_{cc}$  must be applied to this pin thanks to a resistor divider. This voltage value must be slightly higher than 1V 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  $V_{cc}$  threshold level is fixed equal to nearly 17V.

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  $V_{cc}$  because  $V_{cc}$  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,  $P_{in} = 1/2 \times L \times I_{pk}^2 \times f_{osc}$ ).

-MOSFET case:  $(I_{pk})_{max} = 3.5A$

Now, the  $(V_{cs})$  clamp level is nearly 1V (refer to the data sheet)

So,  $(R_s)$  the current sense resistor, must be equal to  $(1V / 3.5A)$ , that is nearly:  $0.28\Omega$  ( $2 \times 0.56\Omega$  in parallel).

-BIPOLAR case:  $(I_{pk})_{max} = 2.5A$

So,  $(R_s)$  must be equal to  $(1V/2.5A)$ , that is:  $0.4\Omega$  ( $3 \times 1.2\Omega$  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 (135W in this application).



6 - Oscillator (pin 10 & 16):

The oscillator frequency is determined by the couple ( $C_T$ ,  $R_{ref}$ ) (refer to the data sheet).

As capacitors have discrete values, the choice of  $R_{ref}$  allows you to fix precisely the oscillator frequency (however,  $R_{ref}$  also fixes the internal current source ( $I_{ref}$ ), which must be lower than  $500\mu A$  and higher than  $100\mu A$ ).

MOSFET case:             $f_{osc} = 50 \text{ kHz} \longrightarrow R_{ref} = 10k\Omega \longrightarrow C_T = 820pF$

BIPOLAR case:             $f_{osc} = 43 \text{ kHz} \longrightarrow R_{ref} = 10k\Omega \longrightarrow C_T = 1nF$

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 15W and a stand-by frequency equal to 20kHz, the calculated  $R_{Fstby}$  and  $R_{Pstby}$  values are:

MOSFET case:             $R_{Pstby} = 10k\Omega$              $R_{Fstby} = 27k\Omega$

BIPOLAR case:             $R_{Pstby} = 10k\Omega$              $R_{Fstby} = 22k\Omega$

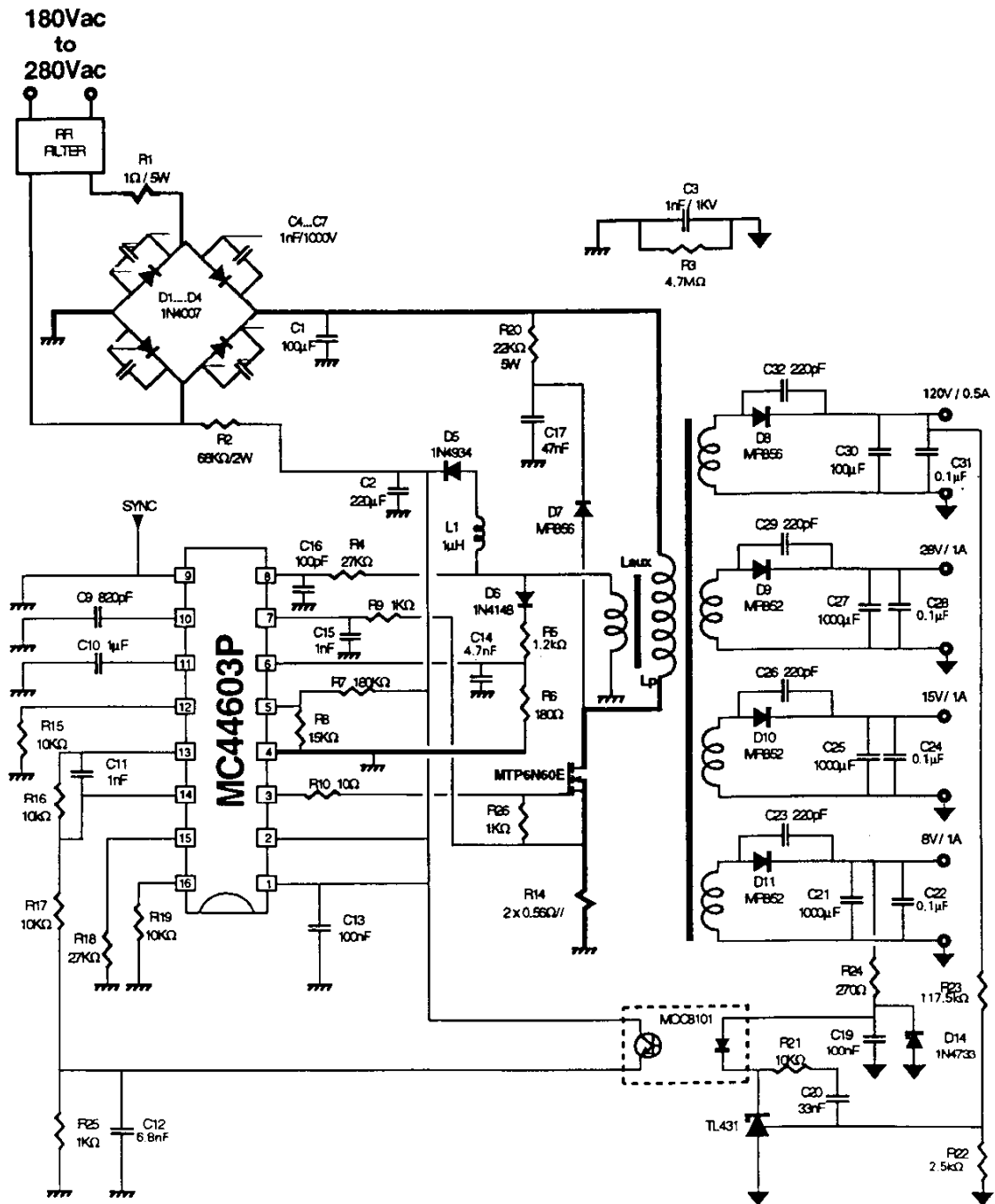


Figure 6. 110W output Off-Line Flyback Converter with MOSFET Switch. 180V - 280V MAINS RANGE

**Table 2.** 110W fly-back converter, 180V - 280V mains range, MC44603 & MTP6N60E

Test	Conditions	Results
Line Regulation	$V_{in} = 180V_{ac}$ to $280V_{ac}$ $F_{mains} = 50HZ$	
	120V $I_{out} = 0.5A$	$\Delta = 0V$
	28V $I_{out} = 1A$	$\Delta = 0V$
	15V $I_{out} = 1A$	$\Delta = 0V$
	8V $I_{out} = 1A$	$\Delta = 0V$
Load Regulation	$V_{in} = 220V_{ac}$ 155V $I_{out} = 0.3A$ to $0.5A$	$\Delta = 0.05V$
Cross Regulation	$V_{in} = 220V_{ac}$ $I_{out} (120V) = 0.5A$ $I_{out} (28V) = 0A$ to $1A$ 120V $I_{out} (15V) = 1A$ $I_{out} (8V) = 1A$	$\Delta = 0V$
Efficiency	$V_{in} = 220V_{ac}$ , $P_o = 110W$	84%
Standby Mode		
P input	$V_{in} = 220V_{ac}$ , $P_{out} = 0W$	3W
Switch. freq.		20KHz fully stable
Output short circuit		Safe on all outputs
Start-up	$P_{in} 110W$	$V_{ac} = 160V$

**MOSFET application: information about the transformer**

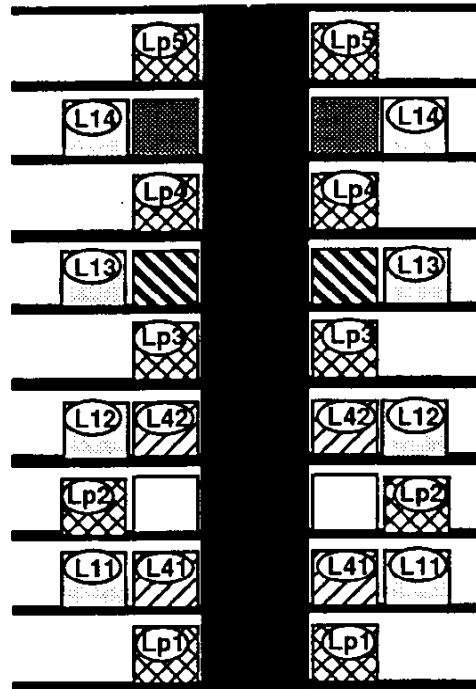
	110W
Lp (turns)	40
Laux (turns)	5
L1 (turns)	40
L2 (turns)	10
L3 (turns)	5
L4 (turns)	3
Al (nH/turns <sup>2</sup> )	274
Core	E-4215A
Material	B2
Former	specific Thomson design
Wire size (mm <sup>2</sup> )	0.315 all windings

**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.



**SMT4**

-  Lp Primary winding  
(Lp1//Lp3//Lp5)+(Lp2//Lp4)
-  Laux Auxilliary winding
-  L1 High voltage secondary winding  
L11//L12//(L13+L14)
-  L2 Secondary winding (28V)  
(2 x 10 turns)
-  L3 Secondary winding (15V)  
(2 x 5 turns)
-  L4 Secondary winding (8V)  
L41//L42

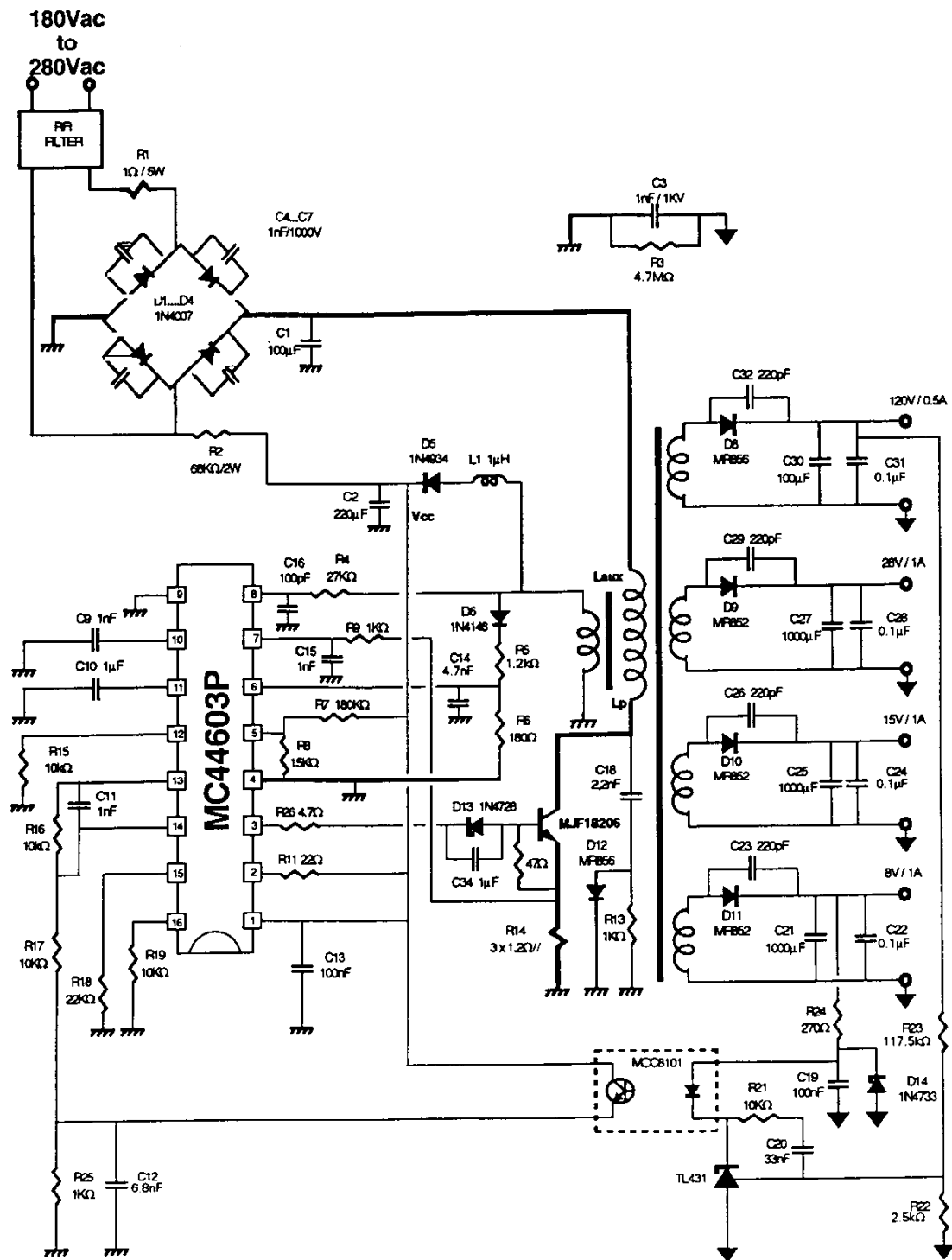


Figure 7. 110W output Off-Line Flyback Converter with Bipolar Switch. 180V - 280V MAINS RANGE.

Table 3. 110W fly-back converter, 180V - 280V mains range, MC44603 MJF18206

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

**BIPOLAR application: information about the transformer**

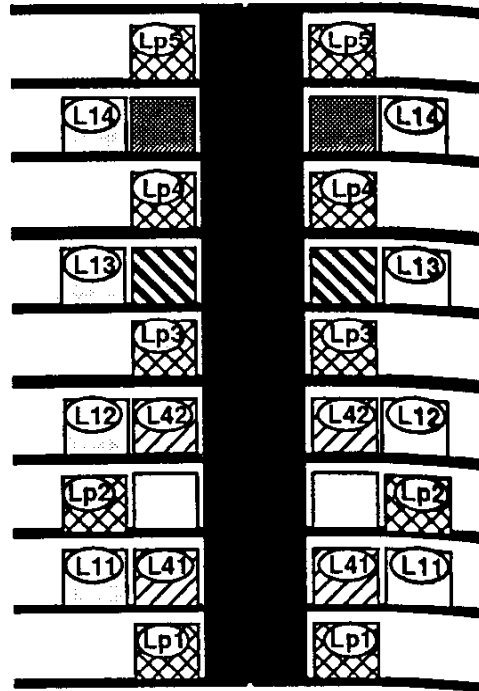
	110W
Lp (turns)	64
Laux (turns)	5
L1 (turns)	40
L2 (turns)	10
L3 (turns)	5
L4 (turns)	3
Al (nH/turns <sup>2</sup> )	250
Core	E-4215A
Material	B2
Former	specific Thomson design
Wire size (mm <sup>2</sup> )	0.315 all windings

**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.



**SMT4**

-  Lp Primary winding  
(Lp1//Lp3//Lp5)+(Lp2//Lp4)
-  Laux Auxilliary winding
-  L1 High voltage secondary winding  
L11//L12//(L13+L14)
-  L2 Secondary winding (28V)  
(2 x 10 turns)
-  L3 Secondary winding (15V)  
(2 x 5 turns)
-  L4 Secondary winding (8V)  
L41//L42

## 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 (135W in our case).

Note that the stand by losses are lower in the 110V 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 110V application. The use of a MOSFET having a lower  $R_{ds(on)}$  (or two MOSFET in parallel) would improve the efficiency.
- because of these losses, the input range of the proposed solution is actually: 90V - 140V

In order to minimise 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 <sub>L</sub>	(ni) @100°C	FERRITE	WIRE Ø (mm)	N <sub>max</sub>
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 12) that will fall within the insulation norms, when using wires whose size is indicated in the "wire" column.

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