HIGH-FREQUENCY POWER TRANSFORMERS

Design a high-frequency power transformer based on flyback topology

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Essentially, switched-mode power supplies, or SMPS, act as AC-to-DC converters. These rectify the AC input voltage (85V-265V AC) to convert it into DC. Depending upon the design considerations, these chop the rectified voltage (DC) at very high frequencies.

SMPS find use in computer power supplies, TV sets, CD players, battery chargers, adaptors, etc. Their major advantages are ligher weight, smaller size, higher efficiency and lesser cost.

Let's now consider the merits of SMPS individually. Lighter weight and smaller size are due to operation at a significantly higher frequency range and use of smaller inductive elements. Rapid switching of the power transistor between saturation and cut-off regions of its operation results in very little energy dissipation and hence reduced heat-sink requirements. Costs are reduced owing to the absence of large bulky power transformers, a huge reduction in volume and power dissipation, smaller material requirements and smaller semiconductor devices.

SMPS have a complex circuitry. Making a traditional 12V DC power supply providing 3A current for a stereo cassette player is not a difficult job for electronic hobbyists with some experience. But designing an SMPS for the same application is quite painstaking.

Power transformer is a crucial part of SMPS. Other components are controllers (PWM ICs), power switches, input/output rectifier and bulk capacitors.

Topology selection

The circuit topology (Fig. 1) has a great impact on the transformer design. Flyback circuits are used primarily at power levels of 0 to 150 watts, forward converters at 50 to 500 watts, half-bridge at 100 to 1000 watts, and full-bridge usually over 500 watts. Full-bridge and half-bridge topologies with full-bridge secondaries have the highest transformer efficiency because the core and the windings are fully utilised.

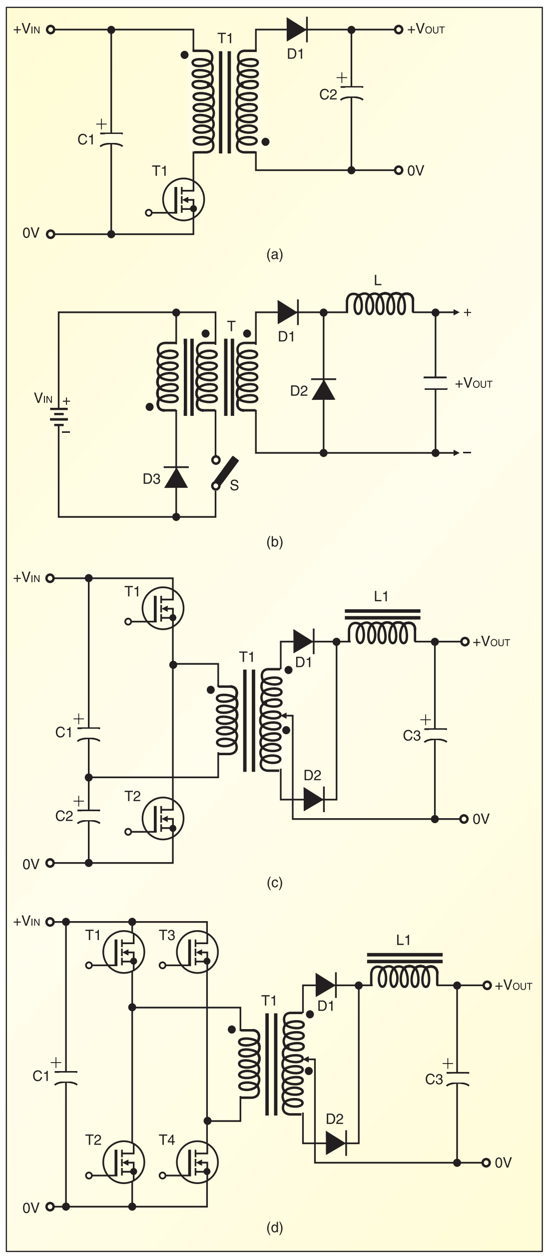


Fig. 1: Various topologies including (a) flyback topology, (b) forward topology, (c) half-bridge topolgy and (d) full-bridge topology

Let's assume that you need 12V DC output at 2A of current for your stereo cassette player from 220V AC, 50 Hz. Since the output power rating is 12V × 2A = 24W, the right topology for this design is the flyback.

To fully understand a flyback power supply design, it is useful to review the theory of flyback topology and the general aspects of a switched-mode power supply such as continuous and discontinuous operation modes of a high-switching frequency transformer design. The power-transformer design is the biggest stumbling block in developing switched-mode power supplies.

In Fig. 1(a) when power switch T1 is 'on' with the application of 'on' pulse from the control circuit (not shown in the figure), the current flows through the primary winding and energy stores within the core. Note that no current can flow through the secondary because of opposite dot polarity (and hence blocked diode D1). When power switch driving pulse from the control circuit is removed (during 'off' time), the polarity reverses and the current flows in the secondary winding. The current flows in either the primary or secondary winding but never in both

TABLE I
Core Size Selection on the Basis of Power Handling Capacity

Output power level	Recommended core types
0-10W	EFD15 SEE16 EF16 EPC17 EE19 EF(D)20 EPC25 EF(D)25
10-20W	EE19 EPC19 EF(D)20 EE or EI22 EF(D)25 EPC25
20-30W	EI25 EF(D)25 EPC25 EPC30 EF(D)30 ETD29 EER28(L)
30-50W	EI28 EER28(L) ETD29 EF(D)30 EER35
50-70W	EER28L ETD34 EER35 ETD39
70-100W	ETD34 EER35 ETD39 EER40 E21

windings simultaneously. Thus the so-called flyback transformer is not a transformer but a coupled inductor.

Discontinuous and continuous modes of operation

A flyback converter has two different modes of operation: discontinuous mode and continuous mode. Both modes require the same circuit; the waveforms of primary and secondary currents through the transformer are shown in Fig. 2. A circuit designed for the discontinuous mode will move into the continuous mode when the output current is increased beyond a certain value. In the discontinuous mode all the energy stored in the primary during the power switch 'on' time is completely transferred to the secondary and to the load before the next cycle,

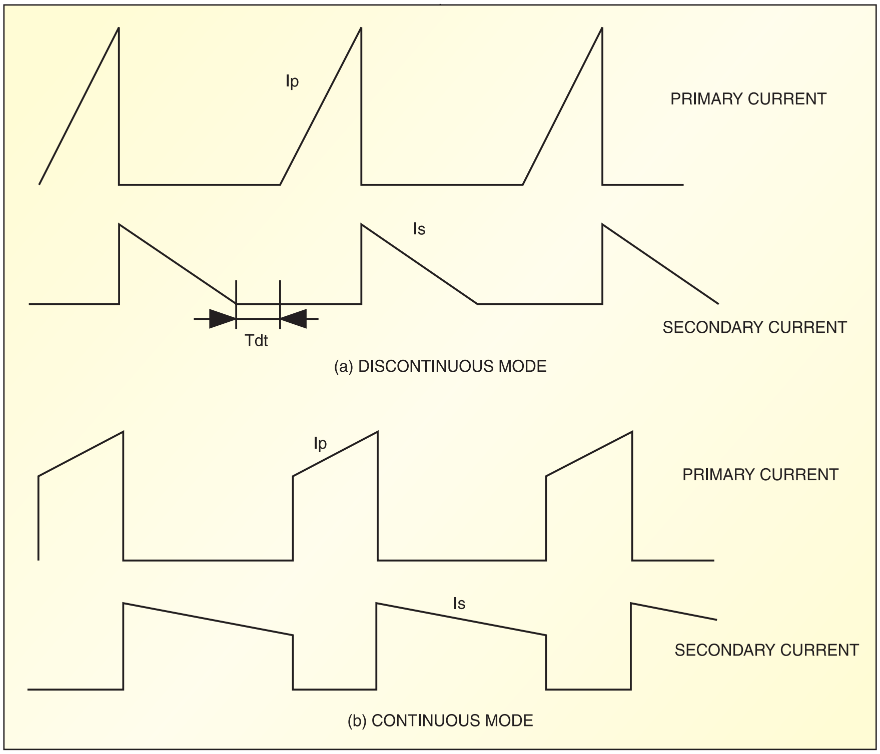


Fig. 2: Primary and secondary currents in (a) discontinuous and (b) continuous mode

and there is also a dead time between the instant the secondary current reaches zero and the start of the next cycle.

In the continuous mode there is still some energy left in the secondary at the beginning of the next cycle. It is possible for flyback to operate in both modes, but it has different characteristics. The discontinuous mode has higher peak currents and therefore higher output voltage spikes during the turn-off. On the other hand, it has faster load transient response and lower primary inductance, and therefore the transformer can be made smaller in size. The reverse recovery time of the output diode is not critical because the forward current is zero before the reverse voltage is applied. Conducted EMI noise is reduced in discontinuous mode because the transistor turns on with zero collector current.

The continuous mode, even if it has lower peak currents, and therefore lower output voltage spikes, is seldom used for low-power applications. Higher voltage and current spikes are not desirable because these exert extra electrical stress on the output diode and the power switch connected in the primary of the transformer.

Flyback transformer design

The flyback topology is used extensively because flyback power supplies require the fewest components. At lower power levels, the total component cost is less than with other techniques. However, between 75 and 150W, increasing voltage and current stresses cause flyback-component cost to increase significantly. At these power levels, topologies with lower voltage- and current-stress levels (such as the forward converter) are more cost-effective, even with higher component counts.

To design a flyback transformer, you need to go through the following steps:

Step 1. Define the power supply parameters pertaining to the transformer design:

(a) Derive output power (Po)

(b) Output voltage (Vo)

(c) Bias voltage (Vb)

(d) AC mains frequency (f_L)

(e) Minimum and maximum AC mains voltage, V_{ACmin} and V_{ACmax}

(f) Maximum duty cycle (Dmax); recommended maximum is 0.5

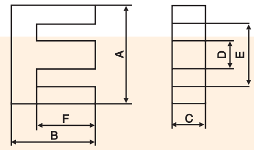
(g) Estimated power supply efficiency (η) at 0.75-0.85

Step 2. Primary inductance calculation:

$$\text{Primary inductance Lp} = \frac{V_{DC\ min} \times Dmax}{Ipp \times fs} \quad \text{....(1)}$$

where $V_{DC\ min} = \sqrt{2} \times V_{AC\ min}$ and Ipp is the peak primary current.

TABLE II
EE and EF Core Specifications



Type	Dimensions (mm)					
	A	B	C	D	E min	F
EE10	10.2 ±0.3	5.5 ±0.2	5.0 +0 −0.3	2.5±0.2	7.5	4.3±0.2
EE10A	10.2 ±0.3	5.5 ±0.2	9.8 +0.3 −0.1	2.5±0.2	7.6	4.3±0.2
EE13	13.0 ±0.4	6.0 +0.2 −0.1	5.9 ±0.3	2.8 +0 −0.4	9.8	4.6 +0.3 −0.1
EE13A	13.0 ±0.4	6.5 ±0.2	9.8 +0.3 −0.1	3.6 ±0.2	8.5	4.6 ±0.2
EE12.6	12.7 ±0.4	6.4 ±0.2	3.3 ±0.2	3.7 ±0.2	8.8	4.65 ±0.3
EE16A	16.0 ±0.4	7.3 ±0.3	5.1 +0 −0.4	4.0 ±0.2	11.7	5.2 +0.2 −0
EE16B	16.0 ±0.4	12.4 ±0.3	5.1 +0 −0.4	4.0 ±0.2	12.4	10.4 ±0.2
EF16	16.0 +0.7 −0.5	8.2 +0 −0.3	4.7 +0 −0.4	4.7 +0 −0.3	11.3	5.7 +0.4 −0
EE19A	19.0 ±0.4	8.0 ±0.3	5.9 +0.1 −0.5	5.0 +0 −0.5	13.8	5.6 +0.4 −0.1
EE19B	19.0 ±0.4	13.6 ±0.3	5.1 +0 −0.5	5.1 +0 −0.5	14.0	11.3 ±0.3
EF20	20.0 ±0.6	9.3 +0 −0.4	5.9 +0.4 −0.5	5.9 +0.4 −0	14.1	6.1 +0.4 −0
EE20	20.5 ±0.5	10.7 ±0.3	7.0 ±0.3	5.0 ±0.3	14.7	7.0 ±0.3
EE20A	20.3 ±0.5	8.4 ±0.3	4.8 ±0.2	4.8 ±0.2	15.3	6.2 ±0.2
EE20/20	20.1 ±0.4	10.15 ±0.25	9.2 ±0.25	4.5 ±0.2	15.2	7.9 ±0.2
EE23	22.7 ±0.3	11.0 ±0.3	10.2 ±0.2	7.5 +0 −0.3	16.4 +0.5 −0	7.5 +0 −0.3
EE25A	25.0 ±0.5	9.9 ±0.3	6.35 ±0.25	6.35 ±0.2	18.6	6.9 ±0.3
EE25B	25.0 ±0.5	9.8 ±0.2	6.1 ±0.3	5.8 ±0.2	18.6	6.8 ±0.2
EE25C	25.4 ±0.6	9.6 +0.3 −0.1	6.75 ±0.2	6.35 ±0.25	18.6	6.5 +0.3 −0.1

$$\text{Average primary current } I_{AV} = \frac{Po}{\eta \times V_{DC\,min}} \quad \ldots.(2)$$

$$Ipp = I_{AV} \times \frac{2}{Dmax}$$

$$= \frac{2Po}{V_{DC\,min} \times \eta \times Dmax} \quad \ldots.(3)$$

Step 3. Calculation of the number of turns in primary, secondary and biasing windings:

The numbers of secondary turns (Np) is calculated as follows:

$$Np = Ns \times \frac{V_{DC\,min}}{Vo + V_D} \times \frac{Dmax}{1\text{–}Dmax} \quad \ldots\ldots\ldots.(4)$$

where V_D is the forward voltage drop of the output diode (D1) as shown in Fig. 2(a) and Ns is the number of secondary turns.

Now let us define turns per volt (Te) to decide Np and Ns. If some conditions (explained later) are not satisfied, we'll have to come back to modify Te.

$$Ns = Te \times Vo \quad \ldots\ldots\ldots\ldots(5)$$

Now from Eq. 4, we can calculate Np.

Step 4. Calculation of the required core size and core air-gap:

Table I explains the power handling capacity of various ferrite cores. After selecting the appropriate core, refer to the manufacturer's datasheet (see Table II) to know the required parameters of the core such as A_L, Ae and Le.

$$A_{LG} = \frac{Lp}{N_p^2} \text{ H/turn}^2$$

Now calculate the maximum flux density Bmax using the effective cross-sectional area for the selected core:

$$Bmax = \frac{Np \times Ip \times A_{LG}}{Ae} \text{ tesla or webber/m}^2$$

The calculated Bmax should be 0.2 to 0.3 tesla. If you get the flux density more than 0.3 tesla, go back to turns per volt (Te). Slightly increase Te to get higher values of Ns and Np and a lower value of A_{LG}. If still you get Bmax more than 0.3 tesla, again increase Te and repeat the process until you get Bmax less than 0.3 tesla.

Now calculate the required air gap, which means you first need to calculate the relative permeability of the ungapped core (μ_r). This is calculated from core parameters Ae (effective cross-sectional area in cm^2), Le (effective magnetic path length in cm^2) and A_L (inductance factor in $nH/turn^2$) as follows:

$$\mu_r = \frac{A_L \times Le}{0.4\pi \times Ae}$$

The gap length (Lg) can now be calculated. The gap should be ground only in the centre leg of the core. If the gap is put into the outer legs, it will need to be half that calculated here. The minimum limit for Lg is 0.051 mm, and Lg is calculated from the following equation:

$$Lg = \left(\frac{0.4\pi \times N_p^2 \times Ae}{Lp} - \frac{Le}{\mu_r}\right) \times 10^{-3} \text{ mm}$$

Step 5. Selection of wire area for primary and secondary windings:

For primary and secondary, choose a wire that doesn't generate too much heat in the winding at the desired current. For that, use the current density (J, in amp/mm^2) to calculate the area of the conductor. The accepted value of J is $3A/mm^2$ to $6A/mm^2$. A good value of J is $4.5A/mm^2$ as this gives a smaller wire size without undesirable temperature rise in the winding and the core.

The area of primary winding conductor:

$$Ap = \frac{\text{Input rms current (Irms)}}{J} \text{ mm}^2$$

Similarly, the area of secondary winding conductor (As):

$$As = \frac{\text{Output current (Io)}}{J} \text{ mm}^2$$

After calculating area of the primary

TABLE III
EE and EF Core Specifications

Type	Ct (mm^{-1})	Ae (mm^2)	Le (mm)	Ve (mm^3)	AL(nH/N^2) ± 25% SK	SP3	SP4	Weight (gm)
EE10	2.1	12.7	26.1	330.7	≥700	810	810	1.8
EE10A	1.1	23.1	26.4	611.6	≥700	810	1000	1.5
EE13	1.7	17.2	30.0	513.8	≥860	1000	1000	3.6
EE13A	0.8	37.4	29.6	1107.6	≥860	1000	1500	3.5
EE12.6	2.4	12.4	29.7	369.0	≥650	900	800	2.0
EE16A	1.92	18.4	35.5	655	≥950	1100	1100	3.3
EE16B	2.8	19.6	55.2	1080	≥680	750	750	5.3
EF16	1.87	20.1	37.6	750	1000	800	800	4.0
EE19A	1.68	23.3	39.2	914	1050	1200	1200	4.6
EE19B	2.65	23.4	62.1	1450	1000	840	840	7.5
EF20	1.2	35.6	42.2	1505	1800	1550	1450	7.8
EE20	1.21	39	47.1	18.50	2100	1900	1800	9.7
EE20A	1.9	22	42.8	942.8	1100	1900	1800	3
EE20/20	1.2	41.1	49.7	2042	2100	2000	1800	5
EE23	0.75	70.0	49.6	3469		2300	2300	17.5
EE25A	1.25	39.6	49.5	1963	≥1600	1900	1900	10.2
EE25B	1.3	36.9	49.7	1823	≥1400	1600	1600	9.0
EE25B/20	1.3	36.9	49.7	1823	≥1400	1600	1600	9.0
EE25C	1.15	43.4	48.2	2091	≥1600	1800	1800	10.0
EF25	1.340	33.5	44.9	1500	≥1630	1900	1900	14.8
EE25D	1.20	40.7	49.2	2004	≥1600	1800	1800	11
EE25E	1.2	49.2	57.9	2850	≥1600	1850	1850	13.5
EE26	0.6	75.8	46	3488.1	≥1600	1850	1700	10
EE28A	0.95	71.6	63.0	4508	3400	3060	2850	35
EE28B	0.57	99.8	48.1	4801	4500	4000	4000	23.5
EE28C	0.82	87.4	73.1	6391	3200	2806	2650	33.5
EE28Q	0.8	83.1	64.4	6350.5	3200	2806	3500	26
EE30	0.5	108.6	57.7	6269.7	2200	1800	1900	20
EE30/30	1.1	59.4	65.5	3879	2200	1900	1900	22
EF32A	0.92	70.1	61.3	4295		2500	2500	21.5
EF32B	0.89	83	74	6140		2400	2400	30
EE33	0.57	124.0	66.1	8192	4500	4000	4000	40.6
EE35A	0.804	86.2	69.3	3973	2990	2600	2600	31
EE35B	0.83	88.3	69.4	6121	3680	3200	3000	31.5
EE40A	0.6	127	77.0	9922	4250	3700	3700	55
EE40B	0.50	153	76.5	11710	4600	4000	4000	60
EE41	0.50	152.0	77.1	11722	4704	4200	4200	61
EE42A	0.7	147.4	96.8	14275	4200	3700	3600	72
EE42B	0.50	184.3	96.8	17843	5400	4700	4700	90
EE42C	0.4	235	97.4	22889	6500	5700	5700	120
EE46	0.6	140.7	84.6	11583		3700	3700	62
EE55	0.35	354	123	43542		6700	6700	216
EE65	0.274	540	147	79000		8000	8000	410

TABLE IV
Wire Parameters

Wire gauge size	SWG diameter (mm)	SWG area (mm^2)
OOOO	10.16	81.032
OOO	9.45	70.102
OO	8.84	61.344
O	8.23	53.170
1	7.62	45.581
2	7.01	38.575
3	6.40	32.154
4	5.89	27.233
5	5.38	22.721
6	4.88	18.694
7	4.47	15.685
8	4.06	12.940
9	3.66	10.516
10	3.25	8.292
11	2.95	6.831
12	2.64	5.471
13	2.34	4.298
14	2.03	3.235
15	1.83	2.629
16	1.63	2.086
17	1.42	1.583
18	1.22	1.168
19	1.02	0.817
20	0.92	0.664
21	0.81	0.515
22	0.71	0.396
23	0.61	0.292
24	0.56	0.246
25	0.51	0.204
26	0.46	0.166
27	0.41	0.132
28	0.38	0.113
29	0.35	0.096
30	0.31	0.073
31	0.29	0.066

and the secondary (in mm^2), select the wire gauge from Table IV.

Switching frequency selection

The operating frequency of the power supply should be selected to obtain the best balance between switching losses, total transformer losses, size and cost of magnetic components and output capacitors.

A high switching frequency reduces the output capacitor value and the inductance of the primary and secondary windings, and therefore the total size of the transformer. But it also increases transformer losses and switching losses of the switch. High losses reduce the overall efficiency of the power supply and increase the size of the heat sink required to dissipate the heat.

Core selection

Ferrite cores are available in many shapes, of which E-core is commonly used in flyback transformers because of its low cost and easy availability. Other types such as EF, EFD, ETD, EER and EI can also be used depending on particular requirements such as height restrictions. RM, toroid and Pot cores are not suitable because of the safety isolation required. EFD are good for low-profile, ETD are good for high-power and EER are good for multiple-output designs.

An example

For a better understanding of the entire power transformer design, here's an example.

Step 1. Power supply specifications:

(a) Output voltage (Vo) = 12V

(b) Output current (Io) = 2A

(c) Total output power Po = $(Vo + V_D) \times Io = (12 + 1) \times 2 = 26W$

Flyback Transformer Application in An SMPS

CD players need 3V to 12V DC to play music and mobile phones need around 4.5V DC to recharge their battery. But we plug the CD player or the cellphone's battery charger into 220V AC, 50Hz through an adaptor (which generally comes with CD player or the cellphone). Now the question is how this 220V AC is converted into 3V or 4.5V DC? Or what is there inside the CD player adaptor or cellphone charger to do this job?

Devices like CD players, cellphone battery chargers, fax machine power supplies, TV power supplies, stereo power supplies, PC power supplies, electronic toy power supplies and PDA adaptors contain nothing but the SMPS shown in Fig. 3. The figure shows the application of a flyback transformer in an SMPS.

The equipment take 220V, 50Hz AC power from the wall-mounted power socket in our home or office. This 220V AC is rectified and filtered into DC voltage without an input isolation transformer. The rectified and filtered DC voltage (in the range of 260V DC to 360V DC) is given to the flyback transformer, where the rectified high-voltage DC is stepped down to 3.3V, 4.5V DC or some other value to feed the device. This is done

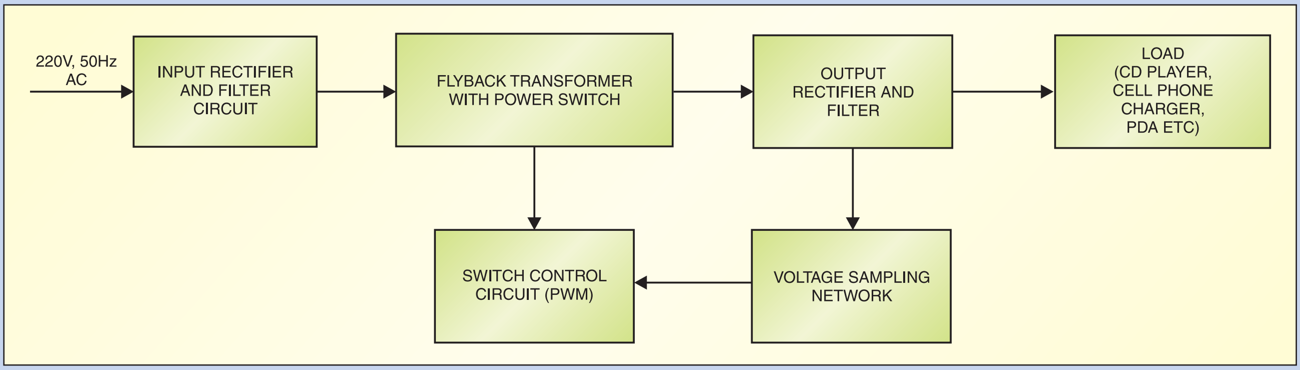


Fig. 3: SMPS block diagram

with the help of a power-switch control pulse-width modulation (PWM) circuit. When you connect the device to the output DC voltage from the flyback transformer after output rectifier, this voltage may vary from the nominal value because of fluctuations in the input AC voltage or overloading.

To maintain the output voltage at the desired level, a sampling network is required. The sampling network gives a sample of the output voltage to check with the reference voltage. For any difference between the sampled voltage and the reference voltage, the PWM circuit takes action to maintain the output voltage constant.

In the block diagram of the SMPS, the power transformer need not be a flyback transformer only. It could be a forward transformer or a half-/ full-bridge transformer. The selection of the topology (type of the transformer) depends on the output power level.

(d) Bias voltage Vb = 18V (generally, 16V to 20V)

(e) AC mains frequency $f_L = 50$ Hz

(f) Minimum AC mains voltage $V_{ACmin} = 85$V and maximum AC mains voltage $V_{ACmax} = 265$V

(g) Maximum duty cycle Dm = 0.5

(h) Estimated power supply efficiency $\eta = 0.85$

(i) Switching frequency fs = 40 kHz

Step 2. Primary inductance calculation:

$$I_{AV} = \frac{Po}{\eta \times V_{DC\,min}}$$

(where $V_{DC\,min} = \sqrt{2} \times 85 = 120$V DC)

$$= \frac{26}{0.85 \times 120} = 0.254\text{A}$$

$$Ipp = I_{AV} \times \frac{2}{Dmax}$$

$$= 0.254 \times \frac{2}{0.45} = 1.128\text{A}$$

$$\text{Primary inductance } Lp = \frac{V_{DC\,min} \times Dmax}{Ipp \times fs}$$

$$= \frac{120 \times 0.45}{1.128 \times 40 \times 10^3} = 1.2 \times 10^{-3}\text{ H}$$

Step 3. Calculation of the number of turns in primary and secondary windings:

Ns = Te × Vo

Taking Te = 1 turn/volt, we get

Ns = 11 × 12 = 12 turns

Now primary turns (Np) are calculated as follows:

$$Np = Ns \times \frac{V_{AC\,min}}{Vo + V_D} \times \frac{Dmax}{1\text{–}Dmax}$$

$$= 12 \times \frac{120}{12 + 1} \times \frac{0.45}{1\text{–}0.45} = 90.62$$

Since turns take a round figure, let's consider Np = 90.

Step 4. Calculation of the required core size and core air-gap:

From Table I, we find that an appropriate core for this design is EE25A. From Tables II and III, core specifications are:

$Ae = 39.6\text{ mm}^2$

$A_L = 1900\text{ nH/turn}^2$

Le = 49.5 mm

$$A_{LG} = \frac{Lp}{N_p^2}\text{ H/turn}^2$$

$$= \frac{1.2 \times 10^{-3}}{90^2} = 1.48 \times 10^{-7}\text{ H/turn}^2$$

Table II shows that the effective cross-sectional area for the selected core (EE25A) is 39.6 mm². Now calculate the maximum flux density as follows:

$$Bmax = \frac{Np \times Ip \times A_{LG}}{Ae}$$

$$= \frac{90 \times 1.128 \times 1.48 \times 10^{-7}}{39.6 \times 10^{-6}}$$

$$= 0.379\text{ tesla}$$

This value of Bmax will saturate the core, hence we have to go back to increase Te and check what happens.

If Te = 1.25, we get:

Ns = 15

Np = 113

$A_{LG} = 9.397 \times 10^{-8}\text{ H/turn}^2$

Now Bmax = 0.302 tesla

As it again shows that the core is operating on the edge of saturation, we calculate Bmax once again with an increased value of Te. If we assume Te = 1.5, we get:

Ns = 23

Np = 173

$A_{LG} = 4 \times 10^{-8}\text{ H/turn}^2$

Bmax = 0.197 tesla

Now, the flux density has become very

low, which means the core is underutilised. Let's calculate once more with Te = 1.35. We get:

Ns = 17

Np = 128

$A_{LG} = 7.324 \times 10^{-8}$ H/turn^2

Bmax = 0.26 tesla

This value is within the acceptable limits.

The relative permeability is calculated as follows (if it is not specified in the core datasheet by the core vendor):

$$\mu_r = \frac{A_L \times Le}{0.4\pi \times Ae}$$

$$= \frac{1900 \times 10^{-9} \times 49.5 \times 10^{-3}}{0.4\pi \times 39.6 \times 10^{-6}}$$

$$= 1.889 \times 10^{-3}$$

Air-gap length (Lg) =

$$= \left(\frac{0.4\pi \times N_p^2 \times Ae}{Lp} - \frac{Le}{\mu_r}\right) \times 10^{-3} \text{ mm}$$

$$= \left(\frac{0.4\pi \times 128^2 \times 39.6 \times 10^{-6}}{1.2 \times 10^{-3}} - \frac{49.5 \times 10^{-3}}{1.889 \times 10^{-3}}\right)$$

The flyback topology is used extensively because flyback power supplies require the fewest components. At lower power levels, the total component cost is less than with other techniques

$\times 10^{-3}$ mm

$= (679.4–26.20) \times 10^{-3} = 0.653$ mm

This gap will be in the centre of the core. The minimum limit for Lg is 0.051 mm.

Step 5. Selection of wire area for primary and secondary windings:

Area of primary winding conductor (Ap):

$$= \frac{\text{Input rms current (Irms)}}{J} \text{ mm}^2$$

$$= \frac{0.436}{4.5} = 0.0970 \text{ mm}^2$$

The rms value of the primary current (Irms) is given by:

$$Ipp \times \sqrt{\frac{Dmax}{3}}$$

$$As = \frac{2}{4.5} = 0.444 \text{ mm}^2$$

After calculating area (in mm^2) of primary and secondary, select the wire gauge from Table IV.

The SWG for primary conductor is 29 and the SWG for secondary conductor is 21 or 20.

The design results can be summarised as follows:

1. Core type: EE25A
2. Primary number of turns (Np) with 28 SWG insulated copper wire = 128
3. Secondary number of turns (Ns) with 21 SWG insulated copper wire = 17
4. Air-gap in the EE core = 0.6 mm

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