

Application Note AN-1024

Flyback Transformer Design for the IRIS40xx Series

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One of the most important factors in the design of a flyback converter power supply is the design of the transformer. The main advantages of the flyback circuit are cost, simplicity and the ease of adding multiple outputs. Flyback topologies are practical and lowest cost for systems up to 100W. Flyback transformer design is a somewhat iterative process, due to the number of variables involved, but it is not difficult, and with a little experience can become a quick and simple process.

APPLICATION NOTE

AN-1024a

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Flyback Transformer Design For The IRIS40xx Series

By Jonathan Adams

TOPICS COVERED

Introduction To Flyback Transformer Design
Power Supply Design Criteria Required
Transformer Design Process
Transformer Construction
Core Types

Wire Table
References
Transformer Component Sources

1) INTRODUCTION TO FLYBACK TRANSFORMER DESIGN

One of the most important factors in the design of a flyback converter power supply is the design of the transformer. Although we call it a transformer it is not actually a true transformer, but more an energy storage device, where during the period of time when the primary switch is on energy is stored in the air gap of the core, and during the off time of the primary switch, this energy is transferred to the outputs. Current flows in either the primary or secondary winding, but not both at the same time. Therefore it can be thought of more as an inductor with secondary windings added.

The main advantages of the flyback circuit are cost, simplicity and the ease of adding multiple outputs. Flyback topologies are practical and lowest cost for systems up to power levels of 100W. Above this power level other methods such as forward converters become more cost effective, due to reduced voltage and current stresses on the devices.

Flyback transformer design is a somewhat iterative process, due to the number of variables involved, but it is not difficult, and with a little experience can become a quick and simple process. Before starting the transformer design it is important to define the power supply parameters such as input voltage, power output, minimum operating frequency, and maximum duty cycle. From there we can calculate the transformer parameters, and select an appropriate core. Iterations may be needed if the calculated parameters do not fall within design guidelines. An Excel spreadsheet is available on the website to simplify the process.

The IRIS40xx series of Integrated Switchers are designed primarily to be used in the quasi-resonant mode which means that the transformer will be operating in a discontinuous mode (The magnetic field is not continuous, it will return to zero when all energy in the transformer is transferred to the secondary side). In PRC mode the transformer will also generally be operating in a discontinuous mode, unless the minimum operating frequency is set very low (about 20kHz which would not generally be practical as this would require a larger core size). So this application note will cover the case for a discontinuous design only.

2) POWER SUPPLY DESIGN CRITERIA REQUIRED

In order to start the design of the transformer some parameters must be defined from the power supply specification. These are:-

- 1) Minimum operating frequency - f_{min}
- 2) Estimated power supply efficiency - $\eta \approx 0.85 \sim 0.9$ (High Vout), $0.75 \sim 0.85$ (Low Vout)
- 3) Minimum DC bus voltage - V_{min} (e.g. 110V for 85Vac minimum input assuming 10V ripple)
- 4) Maximum duty cycle - D_m (recommended maximum is 0.5)
- 5) Value of series resonant capacitor - C_{res} (recommended range is 100pf~1.5nF seen below in Fig1)

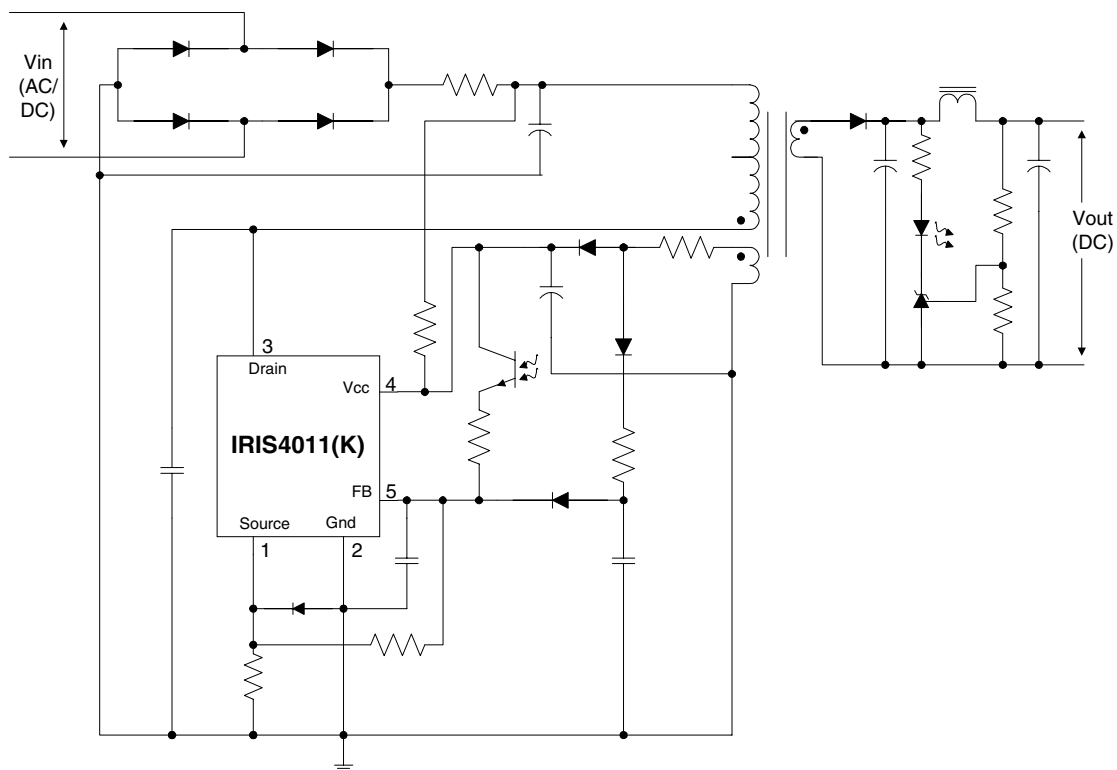


Fig 1) Typical Flyback Power Supply Circuit Using the IRIS40xx Series

3) TRANSFORMER DESIGN PROCESS

The starting point for the design is to calculate the total output power, which is calculated from all the secondary outputs and the bias output including the voltage drop across the output diodes. It is common to use a schottky diode for the main outputs, if the output current is above 1A, or a fast recovery diode if the output is less than 1A, and the bias winding can usually use a simple 1n4148 for the rectifier as this is only a low current output (recommended voltage for the bias is 18V and current is 30mA).

$$P_o = ((V_{O_1} + V_{D_1}) \times I_{O_1}) + \dots + ((V_{O_n} + V_{D_n}) \times I_{O_n}) + ((V_B + V_{D_B}) \times I_B)$$

The output power (P_o) calculated is the total output power.

With this information the primary inductance of the transformer (L_p) can now be calculated from the following equation.

$$L_p = \frac{(V_{\min} \times D_m)^2}{\left(\sqrt{\frac{2 \times P_o \times f_{\min}}{\eta}} + V_{\min} \times \pi \times f_{\min} \times D_m \times \sqrt{C_{res}} \right)^2}$$

The next step is to calculate the required turns ratio for the primary, all secondaries and the bias winding. The following equation will allow you to calculate the primary (N_p) and secondary (N_s) turns.

$$N_p = N_s \times \frac{V_{\min}}{V_o + V_D} \times \frac{D_m}{1 - D_m}$$

Where V_o is the output voltage of the secondary, and V_D is the forward voltage drop of the output rectifier for this secondary. A good starting point is to work on the basis of 1V/turn for the secondary, and calculate the number of primary turns from there. The number of turns for the bias winding N_B is calculated from the following:

$$N_B = N_s \times \frac{V_B}{V_o + V_D}$$

In a power supply with multiple outputs a number of iterations may be needed to find an optimum turns ratio, and some compromise may be needed on the output voltages to ensure the turns are integer values and there are no 1/2 turns.

Now the effective inductance for the gapped core can be calculated. It may be possible to acquire gapped cores with the required A_{LG} value from a core vendor, or a standard core is used and then gapped in the centre leg by grinding to achieve the required A_{LG} . It is calculated from the Primary inductance L_p (in μH) and the number of primary turns (N_p) in the following equation.

$$A_{LG} = 1000 \times \frac{L_p}{N_p^2} \quad \text{nH/turn}^2$$

The average primary current (I_{AV}) can be calculated from the efficiency estimate (η), the required total output power (P_o) and the minimum DC bus voltage (V_{\min}).

$$I_{AV} = \frac{P_o}{\eta \times V_{\min}}$$

The peak primary current (I_p) is now needed and can be calculated from:

$$I_p = I_{AV} \times \frac{2}{D_m}$$

Fig 2) to the left shows the primary current waveform for the discontinuous mode. It shows that during the on time of the switch (t_1) there is a current ramp with the rate controlled by the DC bus voltage and the primary inductance (L_p), ending at a peak current value I_p , which we have just calculated. During the off time (t_2) there is no primary current flowing. The peak flux will occur at the point where $I = I_p$. Due to the quasi-resonant nature of the IRIS40xx circuit, t_1 and t_2 will change depending on the output load and the input voltage. For calculation purposes we use the case of lowest frequency, lowest DC bus voltage, and maximum load as the worst case for the transformer, and therefore the design criteria.

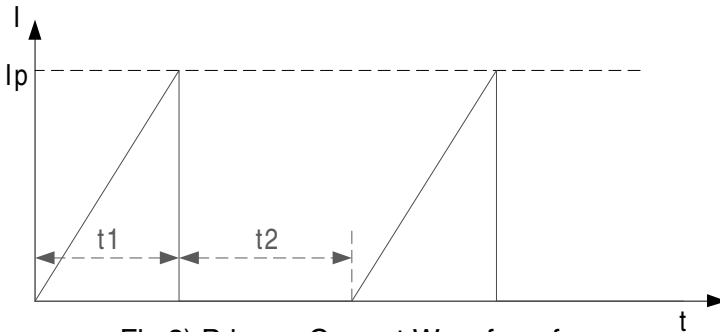


Fig 2) Primary Current Waveform for a Discontinuous Flyback Circuit

RMS primary current (I_{rms}) is needed to be able to calculate the required wires size and is calculated from the following:

$$I_{rms} = I_p \times \sqrt{\frac{D_m}{3}}$$

The next step is to calculate the required core size and air gap. First select a core size, you can use Section 5 which gives an indication of cores types and sizes which could be used and their appropriate wattage levels. Use the following equation to calculate the maximum flux density B_m using the effective cross-sectional area A_e (in cm^2) for the core selected (B_m should be in the range of 2000 to 3000 gauss - below 2000 the core would be underused, and above 3000 there may be a possibility of saturation depending on the ferrite material used).

$$B_m = \frac{N_p \times I_p \times A_{LG}}{10 \times A_e}$$

An alternative is to start with a value for B_m (e.g. 2500) and calculate the minimum A_e needed for the core as below.

$$A_e = \frac{N_p \times I_p \times A_{LG}}{10 \times B_m}$$

B_m can be altered to fall in the required range by changing the number of turns on the secondary (N_s) which directly changes the number of primary turns (N_p). Increasing the number of secondary turns will lower the B_m for a specific core, and conversely lowering the number of secondary turns will increase the B_m .

The AC flux density (B_{AC}) can be used with the core loss curves from manufacturers. This gives the AC component of the magnetic flux instead of the peak to peak. This is simple in the case of a discontinuous transformer design and is simply calculated from:

$$B_{AC} = \frac{B_m}{2}$$

The next step is to calculate the required air gap, which means we first need to calculate the relative permeability of the ungapped core (μ_r). This is calculated from the core parameters, A_e (effective cross-sectional area in cm^2), L_e (effective magnetic path length in cm^2), and A_L (inductance factor in nH/turn^2):

$$\mu_r = \frac{A_L \times L_e}{0.4\pi \times A_e \times 10}$$

The gap length (l_g) can now be calculated. The gap should be ground only in the center leg of the core, this will help to prevent fringing flux from escaping at the edges of the core which could cause EMI noise issues on surrounding components (however for development or small production runs it is acceptable to use insulating material such as tape on the outer legs of the core to give the required gap. It must be remembered though that if the gap is put into the outer legs it will need to be half that calculated here). The minimum limit for l_g is 0.051mm, this is a constraint of the A_{LG} and grinding tolerances. l_g is calculated (in mm) from the following:

$$l_g = \left(\frac{0.4\pi \times N_p^2 \times A_e}{L_p \times 100} - \frac{L_e}{\mu_r} \right) \times 10$$

With core parameters calculated and confirmed we now need to calculate the appropriate wire sizes. First requirement is to calculate the available bobbin width (BW_A) from the actual bobbin width (BW), the number of layers for the primary winding (L) and the margin width (M). The primary winding can be 1, 2 or 3 layers, but minimizing the number of layers will reduce the primary winding capacitance (also insulating the primary winding layers with tape will reduce the winding capacitance significantly) and leakage inductance. The margin size is dependant on the required isolation which is determined by the system input voltage and the appropriate safety agency (details on this can be found in section 4 on transformer construction). Another alternative is to use triple insulated secondary windings which means no margins are required, this method is usually used where size of the transformer is a main criteria, as with this method it is possible to reduce the size of the transformer, but commonly the cost will be increased.

$$BW_A = L \times (BW - (2 \times M))$$

The primary wire size is now calculated based on available winding width, and the number of primary turns to calculate the outside diameter of the wire including insulation (OD in mm). The intention is to make the primary winding cover the entire width of the bobbin, as this will provide the best coupling.

$$OD = \frac{BW_A}{N_p}$$

Using either the wire table in section 5 (which is a good starting point) or a wire table from the appropriate manufacturer it is now possible to select the wire size which matches the calculated OD, and from that reading off the appropriate circular mils (CM) value for the wire size, it is now possible to calculate the current capacity (this is basically the inverse of current density) of the primary winding which is specified as “circular mils per amp” or CMA:

$$CMA_p = \frac{CM}{I_{rms}}$$

The CMA_p calculated should be in the range of 200-500, below 200 the current density is too high, which will lead to heating and power dissipation, and above 500 the wire is underused in terms of current capacity. If the CMA_p calculated is less than 200, iteration is required, this can be done by increasing the number of layers or choosing the next larger core size. If the CMA_p is higher than 500, then reducing the number of layers or selecting a smaller core and recalculating would be the next step. As a guideline the primary wire size should be kept below 26AWG. This is due to the skin effects which cause current to flow near the surface of a wire at high frequencies, thus in a large gauge wire the center of the wire becomes unused, and current crowding occurs near the surface, reducing the effective current carrying area of the wire. This can be overcome by using multifilar windings e.g. multiple strands of 26AWG wire to give the same effective CMA.

Now we need to calculate the wire size for the bias winding and the secondary winding (or windings in the case of a multiple output power supply). To calculate the secondary peak current for the appropriate winding use the following equation :

$$I_{SPx} = \left(I_p \times \frac{N_p}{N_{Sx}} \right) \times \frac{P_{Ox}}{P_o}$$

Where P_{Ox} is the output power for the secondary winding being calculated, and P_o is the total output power calculated at the beginning. This ensures that the peak secondary current calculated is matched with the power for that particular output, this is important for multiple output supplies to ensure the secondary wire gauge is not oversized. This assumes that the secondaries are all individual windings. An alternative is to use stacked secondary windings, which reduces the number of pins required on the bobbin by combining the return connections for the outputs. The two secondary winding schemes can be seen below in Fig 3).

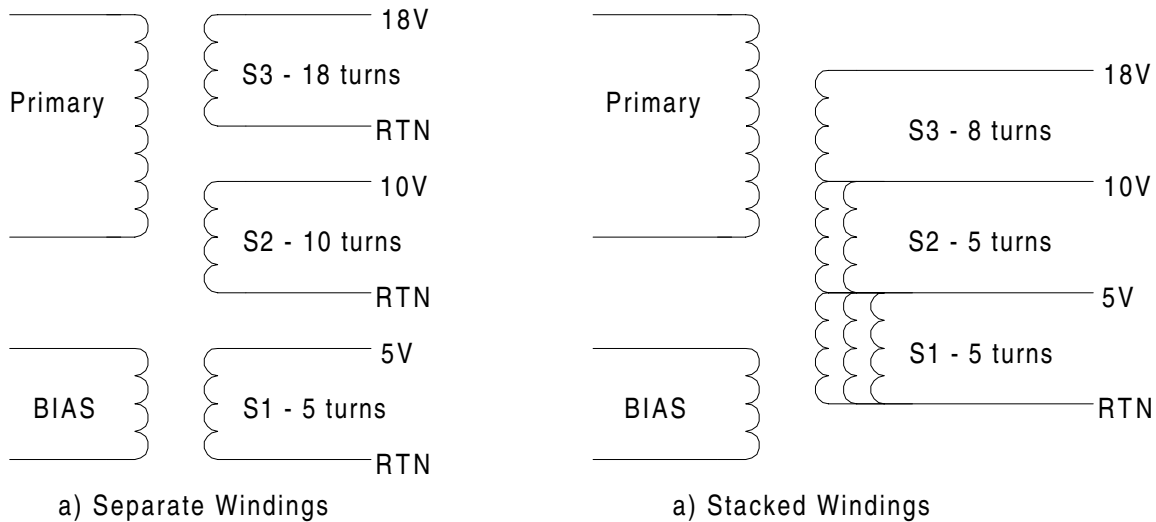


Fig 3) Two Different Secondary Winding Schemes

In the example shown in Fig 3) secondary S1 has to carry the current of S1, S2, and S3 combined, and secondary S2, has to carry the current of S2 and S3 combined, therefore the wire size has to account for this, so the I_{SPx} calculation becomes:

$$I_{SPx} = \left(I_P \times \frac{N_P}{N_{Sx}} \right) \times \frac{\sum P_{Ox}}{P_O}$$

Where $\sum P_{Ox}$ is the sum of the powers for the combined windings e.g. S1+S2+S3 for the S3 winding in scheme in Fig 3b), and S1+S2 for winding S2 in Fig 3b). As S3 is still only carrying its own current, the calculation is simple.

The secondary RMS current (I_{srms}) can now be calculated from:

$$I_{Sxrms} = I_{SxP} \times \sqrt{\frac{1 - D_m}{3}}$$

Fig 4) below shows the IRIS40xx Drain Voltage, the Primary current, the transformer secondary voltage, and the secondary current. This shows the relationship between the primary and secondary , and how the current does not flow in the primary and secondary at the same time.

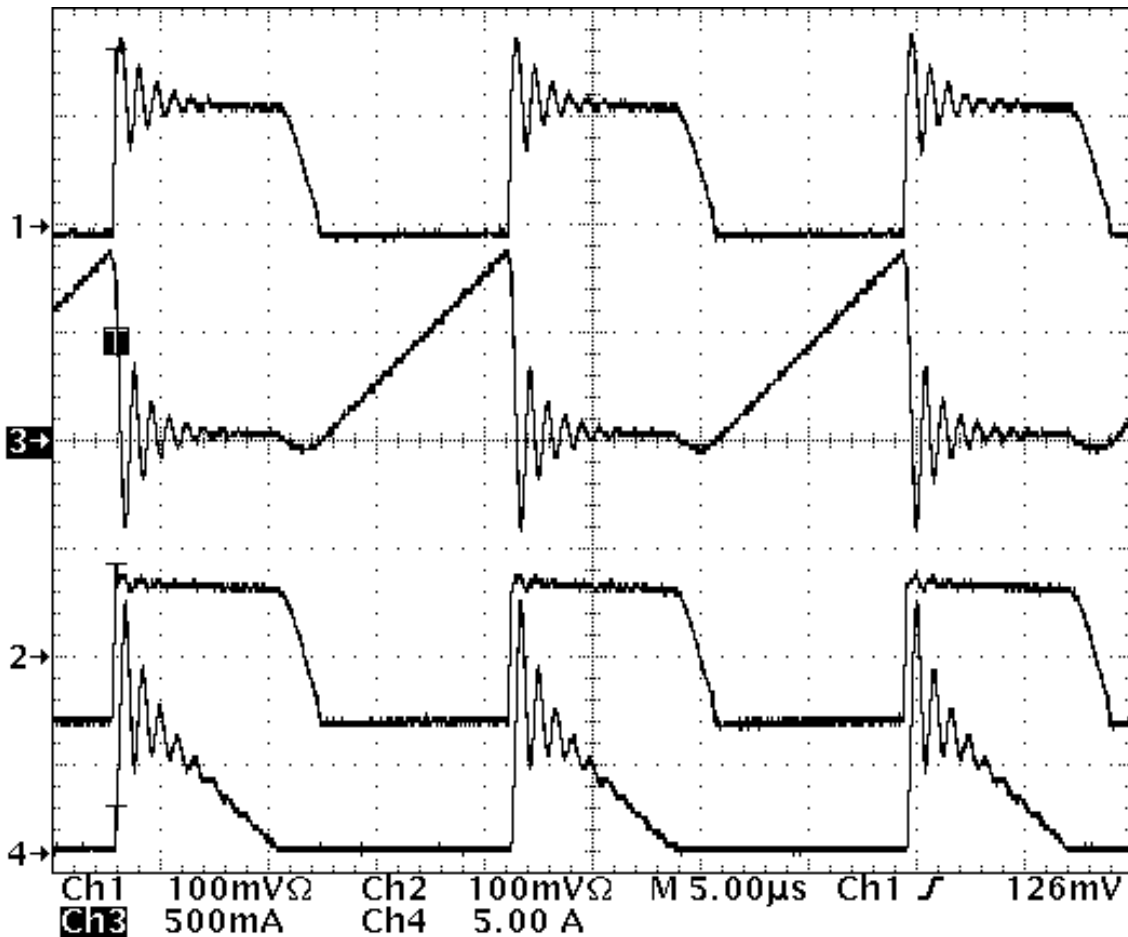


Fig 4) IRIS40xx Drain Voltage(CH1-200V/div), Primary Current(CH3), Secondary Voltage(CH2-20V/div), and Secondary Current(CH4) For a 12V/2A Power Supply at 90Vac in and 1.5A Load

With the secondary RMS current ($I_{S_{rms}}$) calculated it is now possible to calculate the required secondary wire size using:

$$CM_{Sx} = CMA_p \times I_{S_{rms}}$$

Notice that the CMA (current capacity) calculated for the primary is used here to ensure the current capacity is matched for the primary and secondary. Use the CM calculated to select the appropriate wire size from the wire table. Always round down to the next lower AWG number (which is the next larger wire size) where it is reasonably possible. Secondary wire sizes larger than 26 AWG are not recommended for the same reasons mentioned earlier regarding primary wire size, so it may be necessary to use parallel windings of smaller wire size or Litz wire (which is like multifilar wire where the conductors are braided- this type of wire is usually

custom made and can be expensive, but will give the best results). When using parallel conductors make sure the overall CM is within 10% of the value calculated above. Use the same method to calculate the required wire size for the bias winding.

In order to provide the best coupling between primary and secondary, the secondary winding should fill the entire bobbin width. This could be achieved by using parallel windings, as the secondary windings usually only have a small number of turns.

The transformer manufacturer will need the following information to build the transformer:

- Core and Bobbin part number (with required gapped A_L value (A_{LG}))
- The wire gauge and insulation style for each winding
- Safety and Creepage requirements
- Primary Inductance (L_p)
- Number of turns for each winding (N_p , N_B , N_S)
- Pin Connection for the bobbin
- Winding instructions and placement
- Temperature class for operation (e.g. Class A = 105°C)

4) TRANSFORMER CONSTRUCTION

There are 2 main types of transformer design used in flyback transformer construction, these are:

- 1) Margin Wound - As mentioned tape margins are used at the edges of the bobbin to provide the required creepage for the Safety Agencies.
- 2) Triple Insulated - In this case the secondary windings are made with wire which has 3 separate layers of insulation such that any combination of 2 layers will meet the electric strength requirements.

Safety requirements and creepage and electric strength requirements are listed in appropriate standard, one example is for Information Technology Equipment, which is covered by UL1950 in the US, and EN60950(IEC950) in Europe. A creepage distance of 5-6mm is usually sufficient, therefore 2.5mm or 3mm margin tape is usually used on the primary and secondary, in the case of the margin wound construction. Fig 5) shows both the margin wound construction and the triple insulated construction. Margin wound construction is the most common type used. Margin wound construction often ends up being the most cost effective method due to the lower material cost, triple insulated construction will result in a smaller size transformer, as the entire width of the bobbin can be used for winding, because no margins are required, but the material cost is higher and the winding costs may be higher.

Fig 5a) shows the Margin wound construction, in this case the margins are constructed from tape which is cut to the desired width for the margin which is usually 1/2 the required creepage distance (e.g. 3mm for 6mm creepage). The margin tape is then layered to match the height of the windings. The core size selected should allow for the available winding width to be at least twice the creepage distance required to maintain good coupling and minimize the leakage inductance. The primary winding is the first winding on the bobbin, and should be wound so that the start of the winding (the innermost connection to the primary) is the end which will be connected to the Drain pin of the IRIS40xx. This ensures that the point with the maximum voltage swing is shielded by the other windings, therefore minimizing the EMI generated which could couple to other components on the board. If the primary winding is more than one layer, one layer of basic insulation (which is cut to fit between the margins) should be placed between the layers to reduce the possibility of breakdown between layers and also reduce the interwinding capacitance of the primary. Another layer of basic insulation is placed

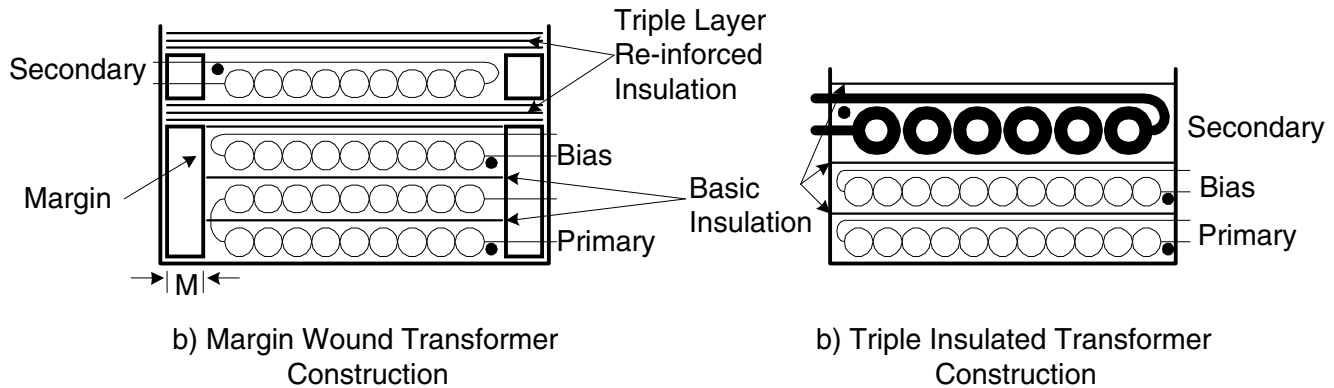


Fig 5) Transformer Construction For Margin Wound and Triple Insulated Transformer Types

over the primary, and the bias winding is placed on top of this. Three layers of tape (which are cut to fit the entire bobbin width) are now placed over the bias winding to meet the requirements for the re-inforced insulation between primary and secondary. Another margin is placed on top of this and the secondary is placed between the margins, so that there is effectively 6mm creepage distance and full voltage isolation between the primary and secondary. Finally another three layers of tape (again the full width of the bobbin) are placed on top of the secondary to hold the windings in place and provide a final re-inforced insulation for the secondary. For each of the windings sleeving must be used with a minimum of 0.41mm wall thickness for the wire which terminates at a pin, to maintain the safety isolation requirements.

Fig 5b) shows the Triple Insulated construction. As can be seen the primary is wound over the full width of the bobbin, then a single layer of tape is put over this before the bias winding. Another layer of tape is placed over the bias winding to prevent damage to the insulation of the triple insulated secondary wire. The secondary winding is now placed on top and secured in place with a final single layer of tape. Care must be taken when winding and soldering to ensure the insulation is not damaged.

4.1) TRANSFORMER MATERIALS

Ferrite Cores

There are a number of manufacturers of ferrite cores which can be used for Flyback transformers. The following materials are suitable for use:

TDK - PC40 or PC44 material
Philips - 3C85, 3C90 or 3F3
Siemens - N27 or N67

There are many core shapes available but the most common used in flyback transformers is an E-core, 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 etc. 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.

Bobbins

The main requirement for the bobbins is to ensure they meet the safety creepage distances, mainly with regards to the primary to secondary pin distance through the core, and the primary pin to secondary winding area distance. Bobbins should be made from materials which can withstand the temperatures encountered during soldering

Insulating Tape

Polyester and Mylar are the most common forms of insulation used usually in tape form, and these can be ordered to the required width for either the basic insulation width or the full width for re-inforced isolation (e.g. 3M #1296 or 1P801 from Lodestone Pacific). The Margin tape is usually a thicker tape so less layers are required, and this is usually a polyester/mat tape such as 3M #44 or 1H860 from Lodestone Pacific.

Magnet Wire

The preferred coating for the magnet wire is a nylon/polyurethane, which will burn off when contacted with molten solder, which allows the transformer to be soldered by dipping into a solder pot. An example is the Belden Beldsol® wire. Normal enameled wire is not recommended as this requires the insulation to be stripped before soldering.

Triple Insulated Wire

In a triple insulated construction, the secondary winding is done using triple insulated wire which is similar to magnet wire in that the main wire is a single core, but with three distinct layers of insulation, such that any of the three combinations of two layers meets the safety requirements for re-inforced insulation.

Sleeving

Sleeving is required for the start and finish of the windings in a margin wound transformer. The sleeving must be recognized by the appropriate safety agencies and needs to have a wall thickness of at least 0.41mm to meet the requirements for re-inforced isolation, teflon or heat shrink tubing is most commonly used due to the heat resistance required, ensuring it does not melt at the soldering temperatures.

Varnish

Varnish is usually used to lock the windings and core in place, preventing audible noise and preventing the ingress of moisture into the transformer. It also helps to improve the voltage withstand capabilities of the insulation, and can improve the thermal conductivity enhancing the transfer of heat from the windings to free air. However this does add a slow step to the process.

4.2) WINDING STYLES

C-winding

This is the most common winding style. Fig 6) shows a C-wound 2 layer primary winding. The C-winding is the easiest to do and also the lowest cost but does lead to higher primary winding capacitance. As can be seen in a two layer C-winding, the primary is wound from one edge of the bobbin to the other and then back again to the start side, which makes for a simple winding process.

Z-winding

Fig 7) Shows a Z-wound 2 layer primary. It can be seen that this method is more complicated than the C-winding and will be more costly to manufacture, but does reduce the winding capacitance.

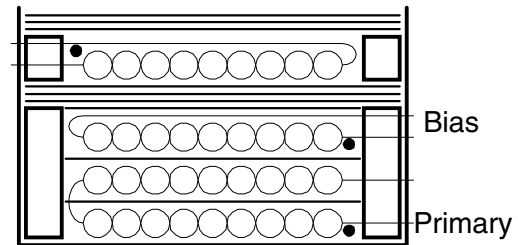


Fig 6) C-wound Primary

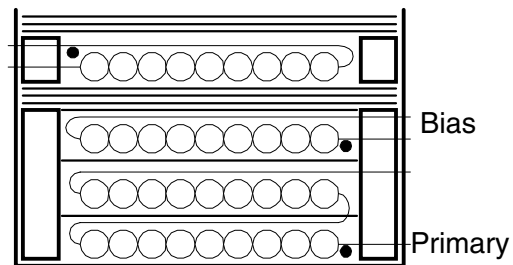


Fig 7) Z-wound Primary

4.3) WINDING ORDER

The Primary winding should always be the innermost winding, this will minimize the mean length per turn, which reduces the primary winding capacitance. By making the Primary winding the innermost winding as previously discussed this means it will be shielded by the other windings, reducing the noise coupling to other components on the board close to the transformer, and by making the start of the winding (the innermost end of the primary) the end which is connected to the drain of the IRIS40xx, this will also reduce noise coupling, as this point (with the maximum voltage swing) is again shielded by the other windings. Using a layer of tape between the layers of the primary winding, this can have a large effect on the primary winding capacitance (reducing it by as much as a factor of four).

The placement of the Bias and Secondary windings is dependant on the type of regulation scheme used. If secondary side regulation is used then the secondary should be the outermost winding, and vice versa if regulation is on the bias winding then the bias winding should be the outermost winding. In a Margin wound design having the secondary winding as the outermost winding reduces the number of margins and layers of re-inforced insulation required. If the bias winding is the outermost winding the coupling to the primary will be less, and coupling to the secondary will be better, improving the output regulation, and also reducing the peak charging effect of the bias supply capacitor by the leakage inductance spike.

4.4) MULTIPLE OUTPUTS

In a multiple output design the highest power level secondary should be the closest to the primary winding to reduce the leakage inductance and ensure the best coupling. The secondary should be wound to fill the entire available winding width, this can be made easier by using multifilar secondaries as discussed earlier,

which also improves the wire usage at high frequencies.

Using the stacked secondaries technique discussed previously will improve the load regulation of the auxilliary outputs, reduce the total number of secondary turns and the number of pins required on the bobbin.

4.5) LEAKAGE INDUCTANCE

Transformer construction has a large effect on the leakage inductance of the primary winding. Leakage inductance leads to the voltage spike at turn off of the mosFET, so minimizing the leakage inductance will result in a lower voltage spike and lower or even no requirement for the snubber circuit on the primary.

Transformer windings should always be concentric i.e. on top of each other to maximize the coupling and therefore reduce the leakage inductance, for this reason split, offset and multi-section bobbins should not be used.

Another technique which can reduce the leakage inductance by half is the use of a split primary winding (fig 8). A split primary winding is one where the first layer of the winding is the inner-most winding and the second layer is wound on the outside. This does require a spare pin on the bobbin to allow for termination of the center point of the primary, but will improve coupling signifi-

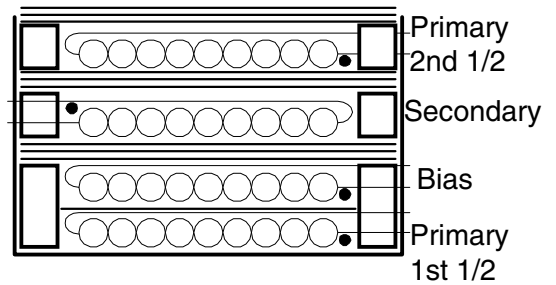


Fig 8) Split Primary Transformer

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5) TRANSFORMER CORE TYPES

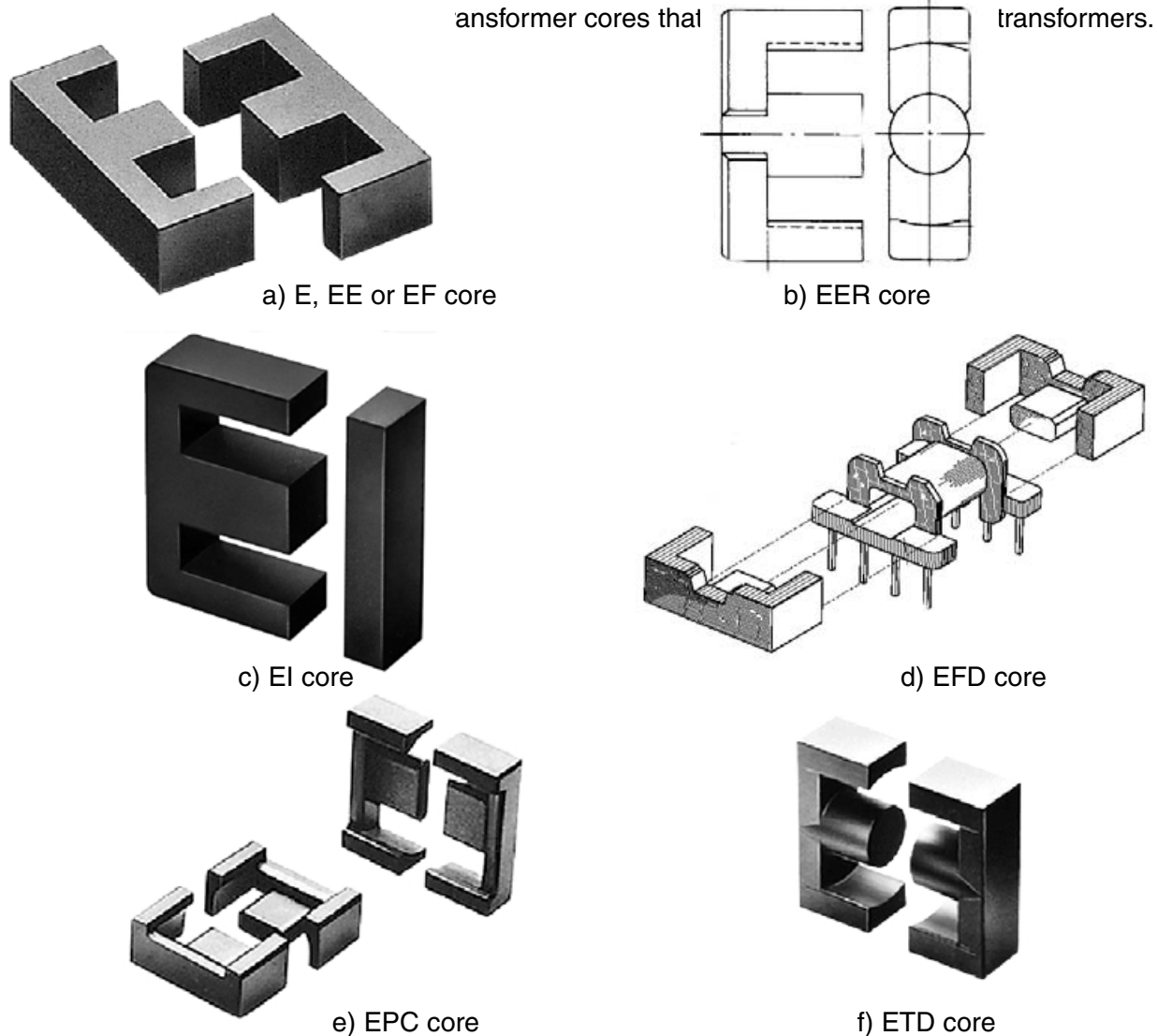


Fig 9)Transformer Core Types For Flyback Power Supplies

The Core type used depends mainly on size restraints. EFD and EPC cores are used when a low profile is required. E, EE, and EF are good general cores and can usually be used with either vertical or horizontal bobbins (vertical bobbins are good when footprint space is at a premium). ETD and EER cores are usually larger, but have a wide winding area, which makes them particularly good for higher power designs or multiple output designs.

Remember a larger core size will be needed for a margin wound type tranformer than for a triple insulated type to allow room for the margins. The core table (table 1) on the next page will help with the core size and

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

Table 1) Transformer Core Table

type selection.

AWG Wire Size	Nearest SWG Wire Size	Nearest Metric Wire Size (mm)	Bare Conductor Area		Outside Diameter (including insulation)	
			cm ² x10 ⁻³	CIR-MIL	cm	inch
14	16	1.60	20.82	4109	0.171	0.0675
15	17	1.40	16.51	3260	0.153	0.0602
16		1.32	13.07	2581	0.137	0.0539
17	18	1.12	13.39	2052	0.122	0.0482
18	19	1.00	8.228	1624	0.109	0.0431
19	20	0.90	6.531	1289	0.0980	0.0386
20	21	0.80	5.188	1024	0.0879	0.0346
21	22	0.71	4.116	812.3	0.0785	0.0309
22		0.63	3.243	640.1	0.0701	0.0276
23	24	0.56	2.588	510.8	0.0632	0.0249
24	25	0.50	2.047	404.0	0.0566	0.0223
25	26	0.45	1.623	320.4	0.0505	0.0199
26		0.40	1.280	252.8	0.0452	0.0178
27	29	0.355	1.021	201.6	0.0409	0.0161
28	30	0.315	0.8046	158.8	0.0366	0.0144
29	31	0.300	0.6470	127.7	0.0330	0.0130
30	33	0.250	0.5067	100.0	0.0294	0.0116
31	34	0.236	0.4013	79.21	0.0267	0.0105
32		0.200	0.3242	64.00	0.0241	0.0095
33		0.180	0.2554	50.41	0.0216	0.0085
34		0.160	0.2011	39.69	0.0191	0.0075
35		0.140	0.1589	31.36	0.0170	0.0067
36	39	0.132	0.1266	25.00	0.0152	0.0060
37	41	0.112	0.1026	20.25	0.0140	0.0055
38	42	0.100	0.08107	16.00	0.0124	0.0049
39	43	0.090	0.06207	12.25	0.0109	0.0043
40	44	0.080	0.04869	9.61	0.0096	0.0038
41	45	0.071	0.03972	7.84	0.00863	0.0034
42	46	0.060	0.03166	6.25	0.00762	0.0030
43	47	0.050	0.02452	4.84	0.00685	0.0027
44			0.0202	4.00	0.00635	0.0025

The wire table will give a good starting point for calculations, but check on the actual outside diameter for the wire being used as insulation thickness varies by manufacturer. The table covers outside diameter for standard single insulated magnet wire, triple insulated wire is not included, check with supplier for details.

7) REFERENCES

- 1) International Rectifier, AN1018a "Using the IRIS40xx Series Integrated Switchers"
- 2) Col. W.T. McLymon, Transformer and Inductor Design Handbook - Second Edition, New York, MarcelDekker Inc. , 1988 ISBN:0-8247-7828-6

8) TRANSFORMER COMPONENT SOURCES

Ferrite Cores

TDK Cores (www.component.tdk.com)
MH&W International, 14 Leighton Place, Mahwah, NJ07430, USA
Tel:1-201-891-8800, Fax:1-201-891-0625, web:www.mhw-intl.com

Philips(www.acm.components.philips.com) Cores
Epcos(www.epcos.com) cores(formerly Siemens Matsushita)
Eastern Components Inc, 21636 Lassen Street, Chatsworth, CA91311, USA
Tel:1-800-824-8596, Fax:1-818-727-0303, web:www.eastern-components.com

Bobbins

The above manufacturers provide matching bobbins for their cores, but bobbins can also be obtained from:

Lodestone Pacific, 4769 Wesley Drive, Anaheim, CA. 92807 USA
Phone: (714) 970-0900, Toll Free: (800) 694-8089, Facsimile: (714) 970-0800
web:www.lodestonepacific.com

Insulating Tape

Lodestone Pacific, 4769 Wesley Drive, Anaheim, CA. 92807 USA
Phone: (714) 970-0900, Toll Free: (800) 694-8089, Facsimile: (714) 970-0800
web:www.lodestonepacific.com

3M Electronic Specialty Markets (ESM)
web:www.3m.com/esm/index.html

Magnet Wire

Belden Wire & Cable, 2200 U.S. 27 South, Richmond, IN 47374
Tel: 1-800-235-3361/1-765-983-5200 Fax: (765) 983-5294 web: www.belden.com

MWS Wire Industries, 31200 Cedar Valley Drive, Westlake Village, CA 91362
Tel: 1-818-991-8553, Fax: 1-818-706-0911, web: www.mwswire.com.

Triple Insulated Wire

Rubadue Wire Co., Inc., 5150 E. LaPalma Avenue • Suite 108, Anaheim Hills • CA 92807
Tel: 1-714-693-5512, Fax: 1-714-693-5515, Web: www.rubaduewire.com