The peak value of the primary current is given by,

$$I_{1+} = \frac{2}{D_{\min}(2-\Delta I_n)}\left[\sum_{i=1}^{K} n_i I_{0i}\right]$$

where n_i is the turns ratio of the ith secondary winding with respect to the primary winding and is given by,

$$n_i = \frac{V_{0i}'}{V_{ccmin}}\left(\frac{1-D_{\max}}{D_{\max}}\right)$$

and

$$\Delta I_1 = \left(\frac{2\Delta I_n}{D_{\min}(2-\Delta I_n)} \sum_{i=1}^{K} n_i I_{0i}\right)$$

The currents flowing in the secondary windings are obtained by reflecting only that part of ΔI_1 to the corresponding secondary winding, which results in the corresponding winding output current, I_{oi}.

$$\text{so } \Delta I_{2i} = \frac{2\Delta I_n}{D_{\min}(2-\Delta I_n)} \cdot I_{0i}$$

where ΔI_{2i} = Secondary current of the ith winding.

All other design equations remain the same.

4

DESIGN OF INDUCTORS

4.1. INTRODUCTION

Even though inductors and transformers are both magnetic components, there is a very important difference in their functioning and design aspect. In a transformer, the core flux (or the flux density) is decided by the magnetizing current. The load current virtually has no say in deciding the core flux (the flux due to the load current is nullified by the counter flux produced by the primary component of the load current). Where as in an inductor, the core flux is decided only by the load current. Thus if the load current increases, there is a possibility that the core may saturate and inductance will come down. So the primary consideration in an inductor is that one has to know the maximum load current and have the core which does not saturate at this current. This can lead to a huge core size if the current to be handled is large. The core size can be reduced considerably by introducing an appropriate air gap in the magnetic circuit. Figure 4.1 (a) and (b) show the effect of an air gap on the B-H characteristics of the magnetic material illustrating clearly that with the airgap, the coil can carry considerably larger current without saturating the core.

4.2 PRINCIPLES OF INDUCTOR DESIGN

There are several approaches to inductor design, two of which are mentioned below :

- Trial and error approach often guided by the "Hanna Curves" [2]
- Area Product approach

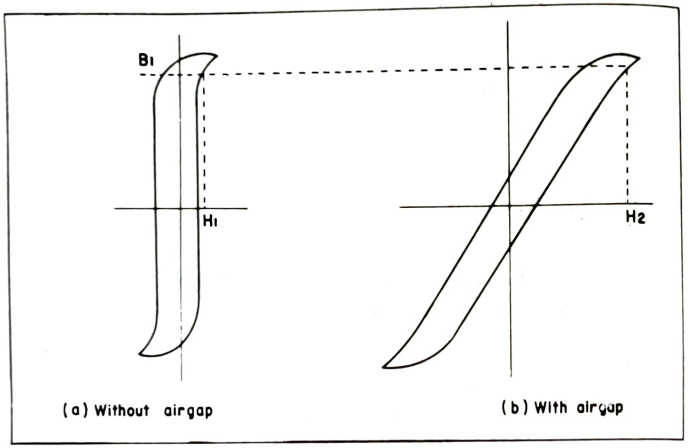

Fig. 4.1 *B-H* Curves

Here the Area Product approach is discussed as it is a sound design technique and is also easy to follow.

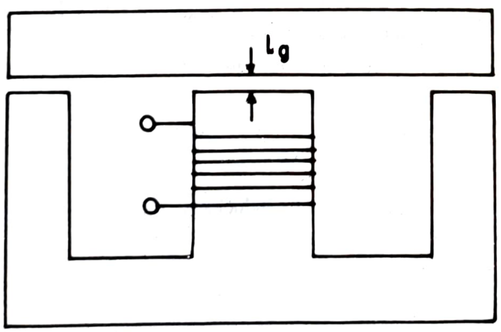

Fig. 4.2 Inductor Geometry with *E-I* core

Figure 4.2 shows the inductor geometry using an *EI* core. Using the *Faraday's law* we have,

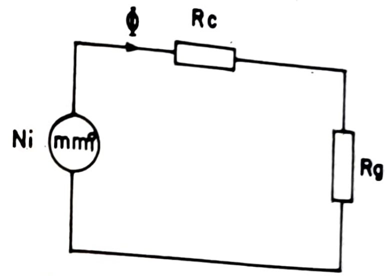

Fig. 4.3 Model of Magnetic circuit for the Inductor

$$e = N\frac{d\phi}{dt} \qquad \ldots(4.1)$$

The equivalent magnetic circuit is shown in figure 4.3. Note that R_c and R_g are the core and air gap reluctances. They are given by the following relationships,

$$R_c = \frac{l_c}{\mu_0\mu_r A_c} \;;\; R_g = \frac{l_g}{\mu_0 A_c}$$

(assuming the area of cross sections of the core and air gap to be equal *i.e.* neglecting the fringing flux.)

Flux in the circuit of the figure 4.3 is given by

$$\phi = \frac{mmf}{R_c + R_g} = \frac{Ni}{\frac{l_c}{\mu_0\mu_r A_c} + \frac{l_g}{\mu_0 A_c}}$$

So the equation 4.1 can be rewritten as,

$$e = N\frac{d\phi}{dt} = \frac{N^2}{\frac{l_c}{\mu_0\mu_r A_c} + \frac{l_g}{\mu_0 A_c}}\frac{di}{dt} = L\frac{di}{dt} \qquad \ldots(4.2)$$

where the inductance L is given by

$$L = \frac{N^2}{\frac{l_c}{\mu_0 \mu_r A_c} + \frac{l_g}{\mu_0 A_c}} \qquad \ldots(4.3)$$

If l_g, the air gap length is zero, then

$$L = \frac{\mu_0 \mu_r A_c N^2}{l_c} \qquad \ldots(4.4)$$

which is the familiar expression of the inductance.

For a core with high permeability, the factor $l_c/(\mu_0 \mu_r A_c)$ can be neglected with respect to $l_g/(\mu_0 A_c)$ *i.e.* the reluctance of the magnetic material is much less than that of the air gap. Thus equation 4.3 simplifies to

$$L = \frac{\mu_0 A_c N^2}{l_g} \qquad \ldots(4.5)$$

i.e.

$$l_g = \frac{\mu_0 A_c N^2}{L} \qquad \ldots(4.6)$$

4.3 DESIGN OF INDUCTOR

Inductor Value : The first step towards the inductor design is to find out the value of L for the particular application. The *Faraday's equation* $e = L\frac{di}{dt}$ is used to find the value of L for any circuit. This equation is best suited for switched mode applications. For circuits based on resonant principle, the L value is determined from the resonant frequency and the Q of the circuit etc.. So depending on the type of application and the configuration of the circuit, the value of L has to be arrived at. In Appendix IV, we shall derive a method for arriving at the value of the inductor for any switching circuit.

Area Product : The energy to be handled by the inductor core is given by

$$E = \frac{1}{2} L I_m^2 \qquad \ldots(4.7)$$

where, E is the energy in joules, L is the inductance in Henrys and I_m is the peak inductor current in amps.

The window area of the core should accommodate 'N' turns of wire cross-section area 'a'. Thus,

$$K_w A_w = N.a \qquad \ldots(4.8)$$

But, $a = \frac{I}{J}$, where I is the rms current through the inductor in amps. and J is the current density in A/mm^2.

So, equation 4.8 can be rewritten as

$$K_w A_w = N\frac{I}{J}$$

Defining crest factor K_c as

$K_c \equiv \frac{I_m}{I} = \frac{(peak)}{(rms)}$, and substituting for I, we have,

$$K_w K_c A_w J = N I_m \qquad \ldots(4.9)$$

From the Faraday's equation, we have

$$e = L\frac{di}{dt} = N\frac{d\phi}{dt} = N A_c \frac{dB}{dt} \quad \text{and}$$

$$L I_m = N A_c B_m \qquad \ldots(4.10)$$

Substituting equation 4.10 in equation 4.7, we have

$$E = \frac{1}{2} N I_m A_c B_m \qquad \ldots(4.11)$$

Substituting for I_m in equation 4.11 from equation 4.9 and rearranging, we have the Area Product for the core given by,

$$A_p = A_w A_c = \frac{2E}{K_w K_c J B_m} \qquad \ldots(4.12)$$

As there is only one winding, K_w can be chosen as 0.6.

The core can be chosen by comparing the area product value obtained from equation 4.12 with the Appendix-I.

No. of turns : The number of turns can be calculated from equation 4.10, which can be rewritten as

$$N = \frac{LI_m}{A_c B_m} \qquad \ldots(4.13)$$

Gauge of wire : The cross-section area of the wire can be calculated from the formula,

$$a = I/J \qquad \ldots(4.14)$$

The gauge of the wire can be decided by comparing the calculated wire cross section from equation 4.14 with Appendix-II.

Air gap, *lg* : As per section 4.1, air gap, l_g, is used to reduce the core size.

From the Faraday's equation, we have,

$$e = N\frac{d\phi}{dt} = L\frac{di}{dt}, \text{ we have}$$

$$B = \frac{LI}{NA_c} \qquad \ldots(4.15)$$

where $B = \dfrac{\phi}{A_c}$

from Ampere's Law, we have

$$mmf = NI = \int H.dl, \text{ which gives rise to}$$

$$H = \frac{NI}{l_m} \quad \text{where } l_m \text{ is the mean magnetic path length}$$

$$\text{so, } B = \mu\frac{NI}{l_m} \qquad \ldots(4.16)$$

Equating equations 4.15 and 4.16, we have

$$\frac{l_m}{\mu A_c} = \frac{N^2}{L} \qquad \ldots(4.17)$$

$\dfrac{l_m}{\mu A_c}$ is the reluctance of the magnetic path which can be split into $\dfrac{l_c}{\mu_0 \mu_r A_c} + \dfrac{l_g}{\mu_0 A_c}$,

where,

l_c is the mean magnetic path in the magnetic material, m

l_g is the air gap length, m

A_c is the core section, m^2

If the material is of high permeability one, then the reluctance is contributed mainly by the air gap, so

$$\frac{l_c}{\mu_0 \mu_r A_c} \ll \frac{l_g}{\mu_0 A_c}$$

so equation 4.17 can be rewritten as,

$$l_g = \frac{\mu_0 N^2 A_c}{L} \qquad \ldots(4.18)$$

In calculating N, B_m is assumed, which may not be the exact B_m in the core, so the air gap calculated may not be exact. So in many cases, value of L may have to be trimmed by slightly adjusting the airgap.

4.3.1 Summary of Design Procedure

- **Determine L for the particular application.**
- **Calculate Area Product**

$$E = \frac{1}{2}LI_m^2$$

$$A_p = A_w A_c = \frac{2E}{K_w K_c J B_m}$$

take $B_m = 0.2T$ for ferrite, $B_m = 1T$ for CRGO, $J = 3$ A/mm^2 and $K_w = 0.6$.

Choose the core from Appendix-I.

- **Calculate the number of turns**

$$N = \frac{LI_m}{A_c B_m}$$

- **Determine the Gauge of the wire**

$$a = I/J$$

Choose the Gauge of the wire from Appendix-II.

- **Cross Check**

The inequality, $A_w K_w > a.N$ should be satisfied, or else repeat the calculations for the number of turns and guage of wire after choosing the next bigger core.

Note that the value of 'a' should be the actual cross section area of the wire used and not the calculated value.

Calculate air gap length, l_g

$$l_g = \frac{\mu_0 N^2 A_c}{L}$$

EXAMPLE 4.1

Design an Inductor for a Buck Converter Configuration as shown in figure 4.4, for the following specifications :

Output voltage, V_0	5V
Output current, I_0	5A
Switching Freq., f_s	40 KHz
Input Voltage, V_g	12 V ± 10%

Solution :

Determine L : The L for this Converter is given by

$$L = \frac{V_0(1 - D_{min})}{\Delta I f_s}$$

ΔI is the current ripple in the inductor. Usually 10 to 25% of I_0 is taken as ΔI. Here ΔI = 10% of I_0 is taken.

where $D_{min} = \dfrac{V_0}{V_{gmax}}$

Substituting the appropriate values in the above equations, we find that

$L = 0.1553$ mH

Area Product : The energy and area product calculations are as follows

$$E = \frac{1}{2} L I_m^2$$

where $$I_m = I_0 + \frac{\Delta I}{2}$$

Substituting the values in the above equations, we have

$$E = 2.14 \times 10^{-3} \text{ joules}$$

$$A_p = A_w A_c = \frac{2E}{K_w K_c J B_m}$$

take $B_m = 0.2$T for ferrite, J = 3A/mm^2 (3×10^6 A/m^2), $K_c = 1$ (for square wave) and $K_w = 0.6$.

Substituting the values in the area product equation, we have

$A_p = 1.18888 \times 10^{-8}$ m^4 = 11888.8 mm^4

Now choose the core from Appendix-I which has a A_p higher than the value calculated above

P 36/22 is proper choice - (A_c = 201 mm^2, A_w = 101 mm^2, A_p = 20100 mm^4)

No. of turns: The equation for the number of turns is given by

$$N = \frac{LI_m}{A_c B_m}$$

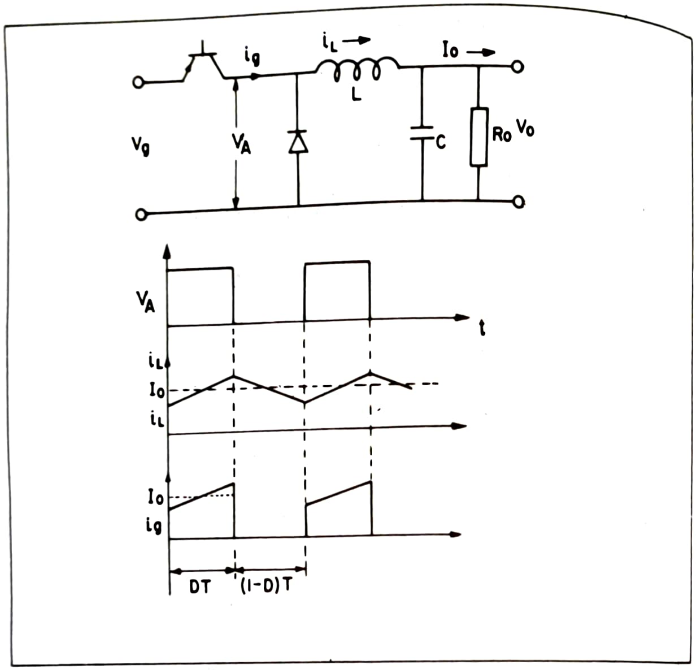

Fig. 4.4 Buck Converter and relevant waveforms

Substituting the values for the variables, we have

N=21 turns (taking the next higher integer if the calculation does not give an integer value).

Wire gauge : The gauge of the wire can be calculated from the equation given below, taking

$$J = 3 \text{ A/mm}^2$$

$$a = \frac{I}{J}, \quad \text{where } I = I_0$$

Substituting the values of the variables in the above equation, we have,

$$a = 1.6666 \text{ mm}^2$$

Now choose the wire gauge from Appendix-II, which has a cross section area greater than the value calculated above.

SWG 16 is a proper choice ($a = 2.075 \text{ mm}^2$)

Cross check : The inequality $A_w K_w > aN$ has to be checked.

$$\text{So } A_w K_w = 101 \times 0.6 = 60.60 \text{ mm}^2$$

and $a\,N = 2.075 \times 21 = 43.57 \text{ mm}^2$

So the inequality is satisfied, which means that the windings will fit into the available window area.

Air gap Length : The air gap length, l_g, is given by the equation,

$$l_g = \frac{\mu_0 N^2 A_c}{L}, \text{ where } \mu_0 = 4\pi \times 10^{-7} \text{ H/m}.$$

Substituting the variable values to the above equation, we have,

$$l_g = 0.717 \times 10^{-3}\, m = 0.717 \text{ mm}$$

4.4. SOME REMARKS ON INDUCTORS

The following remarks are to be noted carefully.

As the permeability of the core is not a constant parameter and the energy stored on the core material cannot be totally neglected, the above design when implemented may not give the exact value of required L. However this design procedure is satisfactory for most of the cases. The actual value of desired L may have to be trimmed by adjusting the air gap.

- The assembly of the core is done in a slightly different manner than that of the transformer, when laminations are used to make the inductor. In the case of the inductor, all E laminations are put together and all I laminations are held together and air gap is put between the E laminations and the I laminations. Thin Hylem/leatheroid/fibre-glass/melinex can be used as air gap. Now the assembly is tightened by means of clamps.
- In the case of potcores, EE and EI cores, a gap of $l_g/2$ only need be introduced in the centre to achieve effective air gap of l_g, due to the geometry of the core.
- The inductor fabricated is not a pure L but has a resistance also. Its value can be computed using the same procedure used in the case of transformer. If there is a particular constraint on R (e.g. Low copper loss requirement), then the wire size has to be chosen properly. One should not forget to cross check the space requirement in case the wire size is altered. Due to the copper losses, the actual energy handled by the core is slightly higher than that given by equation 4.7. An efficiency factor can be incorporated in equation 4.12 so that the required A_p is slightly higher.
- When using *ferrite cores*, once the core is chosen, calculation of the number of turns is often simplified by using the so called A_L value of the core. A_L value represents the inductance factor. The formula for the number of turns is given by,

$$L = A_L N^2$$

When air gap is to be put, a ferrite core of pre-adjusted airgap can be chosen and its A_L value can be used in the design [2].

4.5. NON LINEAR INDUCTORS

Generally in Switched Mode Power Converters, the inductors operate in the linear region. The closed loop dynamics is controlled by L & C values. In addition the L decides the mode of operation (continuous/discontinuous). From the first consideration, the value of L should be small. From the second consideration, the value of L should be large enough, so that it does not get into discontinuous conduction mode. In addition, this compromise has to hold good at all loads. Generally at light loads large

L is required and small L at higher loads, to get a better dynamic performance at higher loads. These two requirements cannot be met by a linear inductor. Hence non linear inductors are preferred in such cases. Inductor characteristics is shown in figure 4.5.

The core cross section at the centre limb is cut as shown in the figure 4.5 (b). At low currents the inductance offered will be high, but as the current increases, the bottleneck at the centre limb starts going towards saturation and therefore lowers the inductance at higher currents, thereby giving an effect of dynamic air gap change. By changing the core section in the centre limb, different L-I characteristics can be obtained. This way the output performance of the SMPCs can be improved.

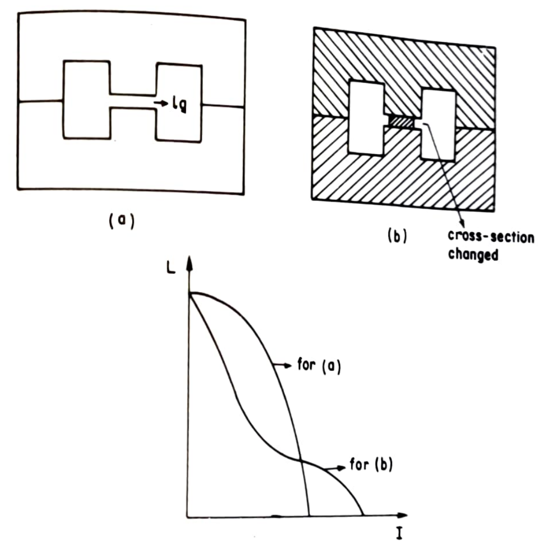

Fig. 4.5 Non Linear Chokes

REFERENCES :

1. N. Radhakrishnan and S. R. Bhat; "Design & Technology of Low Power Transformers and Inductors", CEDT publication, July 1988.
2. M. A. Nadkarni and S. R. Bhat, "Pulse Transormers : Design and Fabrication" CEDT series, Tata McGraw Hill Book Co., 1985

5

DESIGN OF CURRENT TRANSFORMERS

Current transformers (CTs) are mainly used to measure the current flowing in high power circuits for use in feedback circuits, measurement purposes etc.

Alternative way of measuring currents is by using current shunts (small resistor), put in series with the current path to be measured. However, current transformer has the following advantages :

- It provides isolation between the high power circuit and the measuring or control circuit.
- It provides good common mode rejection, e.g. in current mode control application.
- It has less power losses than the current shunts.

In switch mode power converters, one encounters two types of currents. For example, in forward converter, the current flows only in one direction where as in Half, Full bridge converters, the current flow is bidirectional. Design of CTs suitable for both types of applications will be covered in this chapter.

5.1 PRINCIPLES OF BIDIRECTIONAL CT DESIGN : [1]

While the basic principle of operation is similar to that of voltage transformer, there are many peculiar features and constraints in designing a current transformer (*CT*). Figure 5.1 (a) shows a *CT* configuration and figure 5.1 (b) shows an equivalent circuit of *CT*, and for bidirectional current flow.

APPENDIX - I

Physical, Electrical and Magnetic characteristics of ferrite cores

CORES without air gap	mean length per turn l_t mm	mean magnetic length l_m mm	core cross section area $A_c \times$ 100 mm^2	window area A_w $\times 100$ mm^2	area product $A_p \times 10^4$ mm^4	effective relative permeability $\mu_r \pm 25\%$	A_L nH/turns^2 ±25%
POTCORES - CEL HP$_3$C grade, (*Philip 3B7 grade)							
P 18/11	35.6	26	0.43	0.266	0.114	1480	3122
P 26/16	52	37.5	0.94	0.53	0.498	1670	5247
P 30/19	60	45.2	1.36	0.747	1.016	1760	6703
P 36/22	73	53.2	2.01	1.01	2.010	2030*	9500*
P 42/29	86	68.6	2.64	1.81	4.778	2120*	10250*
P 66/56	130	123	7.15	5.18	37.03		
EE - CORES - CEL HP$_3$C grade							
E 20/10/5	38	42.8	0.31	0.478	0.149	1770	1624
E 25/9/6	51.2	48.8	0.40	0.78	0.312	1840	1895
E 25/13/7	52	57.5	0.55	0.87	0.478	1900	2285
E 30/15/7	56	66.9	0.597	1.19	0.71		
E 36/18/11	70.6	78.0	1.31	1.41	1.847	2000	4200
E 42/21/9	77.6	108.5	1.07	2.56	2.739	2100	2613
E 42/21/15	93	97.2	1.82	2.56	4.659	2030	4778
E 42/21/20	99	98.0	2.35	2.56	6.016	2058	6231
E 65/32/13	150	146.3	2.66	5.37	14.284	2115	4833

CORES without air gap	mean length per turn l_t mm	mean magnetic length l_m mm	core cross section area $A_c \times 100$ mm^2	window area A_w $\times 100$ mm^2	area product $A_p \times 10^4$ mm^4	effective relative permeability $\mu_r \pm 25\%$	A_L nH/turns^2 ± 25%
UU - CORES							
UU 15	44	48	0.32	0.59	1.190		1100
UU 21	55	68	0.55	1.01	0.555		1425
UU 23	64	74	0.61	1.36	0.823		1425
UU 60	183	184	1.96	11.65	22.83		1900
UU 100	29.3	308	6.45	29.14	187.95		3325
TOROIDS - CEL HP$_3$C							
T 10	12.8	23.55	0.062	0.196	0.012	2300	765
T 12	19.2	30.40	0.12	0.442	0.053	2300	1180
T 16	24.2	38.70	0.20	0.785	0.157	2300	1482
T 20	25.2	47.30	0.22	0.950	0.213	2300	1130
T 27	34.1	65.94	0.42	1.651	0.698	2300	1851
T 32	39.6	73.00	0.61	1.651	1.010	2300	2427
T 45	54.7	114.50	0.93	6.157	5.756	2300	2367

APPENDIX - II

Wire Size Table

SWG	Dia with enamel mm	Area of bare conductor mm^2	R/Km @20°C ohms	Weight Kg/km
45*	0.086	0.003973	4340	0.0369
44	0.097	0.005189	3323	0.0481
43	0.109	0.006567	2626	0.0610
42	0.119	0.008107	2127	0.0750
41	0.132	0.009810	1758	0.0908
40*	0.142	0.011675	1477	0.1079
39	0.152	0.013700	1258	0.1262
38*	0.175	0.018240	945.2	0.1679
37	0.198	0.023430	735.9	0.2202
36	0.218	0.029270	589.1	0.2686
35*	0.241	0.035750	482.2	0.3281
34	0.264	0.04289	402.0	0.3932
33	0.287	0.05067	340.3	0.4650
32*	0.307	0.05910	291.7	0.5408
31	0.330	0.06818	252.9	0.6245
30	0.351	0.07791	221.3	0.7121
29*	0.384	0.09372	184.0	0.8559
28	0.417	0.11100	155.3	1.0140
27	0.462	0.13630	126.5	1.2450
26*	0.505	0.16420	105.0	1.4990
25	0.561	0.20270	85.1	1.8510
24*	0.612	0.24520	70.3	2.2330
23	0.665	0.29190	59.1	2.6550
22*	0.770	0.39730	43.4	3.6070
21	0.874	0.51890	33.2	4.7020
20*	0.978	0.65670	26.3	5.9390
19	1.082	0.81070	21.3	7.3240
18*	1.293	1.16700	14.8	10.5370
17	1.501	1.58900	10.8	14.3130
16	1.709	2.07500	8.3	18.6780
15	1.920	2.62700	6.6	23.6400
14*	2.129	3.24300	5.3	29.1500
13	2.441	4.28900	4.0	38.5600
12	2.756	5.48000	3.1	49.2200
11	3.068	6.81800	2.5	61.0000
10	3.383	8.30200	2.1	74.0000
9	3.800	10.5100	1.6	94.0000
8	4.219	12.9700	1.3	116.0000

*Standardised at CEDT

APPENDIX - IV

INDUCTOR DESIGN SUPPLEMENT

Inductors are energy storage elements, which stores the energy from the system source when the source is active and eventually delivers the energy back to the system when the source is inactive. A source is said to be active when it is delivering power, and inactive when it is either not delivering power or sinking power.

Inductors are generally used wherever there is a requirement to smoothen the current flow. In most switching applications, inductors are invariably used in the output filters to smoothen the current flow through the load.

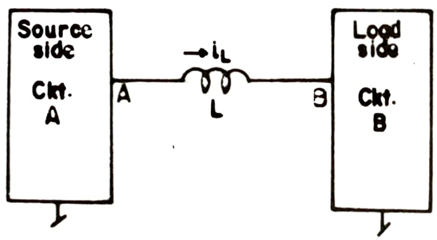

Fig. A4.1 :Inductor application topology

Consider an inductor as shown in figure 1. Let point A be connected to the source side of the inductor and point B be connected to the load side of the inductor.

Let the source be active for a duration DT and the duration for which the source is inactive is $(1 - D)T$ where T is the period of action and inaction of the source *i.e.* $1/T = f$ is the frequency at which the source becomes active and inactive, and D is the duty cycle.

When the source is active during duration DT, the energy is being stored in the inductor, and the direction of current is as shown in figure 1. Point A of the inductor is positive w.r.t to point B. When the source goes inactive, the energy stored in the inductor is released to the load. Now the inductor acts as a generator and hence point B is at a higher potential than point A. It is assumed that the source side and load side circuits of

figure 1 are designed such that there is a path to maintain the current, i_L, through the inductor, even when the source is inactive, in the direction as shown in figure 1. It is also interesting to note that the average voltage across an inductor is always zero.

DETERMINING VALUE OF INDUCTORS FOR SMPCs

In all switched mode converters, the inductor will be used as shown in figure 2. When the source is active the voltage, V_{AC} is a pulsed waveform with duty cycle D. When the source is inactive, there exists a free wheeling path in the source side circuit which causes V_{AC} to become zero. The above configuration applies for forward converter, half bridge, full bridge and pushpull converters.

So $V_{AB} = V_{AC} - V_{BC}$

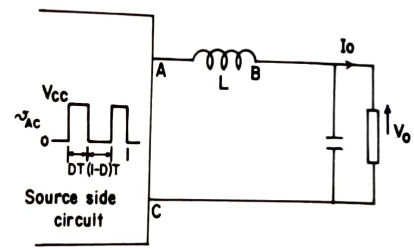

Fig. A 4.2 :Inductors for SMPCs

During the period DT, when the source is active,

$$V_{AB} = V_{CC} - V_0 = V_L$$

and during the period (1 – D) T, when the source is inactive

$$V_{AB} = -V_0 = V_L \quad V_{AC} = 0$$

or $\quad V_{BA} = V_0$

Now from Faraday's Law

$$V = L\frac{di}{dt} \quad \text{we have}$$

$$V_{ac} - V_0 = L\frac{\Delta I}{DT} \quad \text{or} \quad V_0 = L\frac{\Delta I}{(1-D)T}$$

Either of the above equations can be used to obtain the L value, so from the 2nd equation we have,

$$L = \frac{V_0(1-D)T}{\Delta I}$$

where ΔI is the ripple in the inductor current. It is generally in the order of 1% to 10% of I_0.

It should be noted here that T is not the switching period T_S of the converter. For the forward converter, T is same as the switching period T_S, but for half bridge, full bridge and push pull converters T is half of T_S because of the ORing effect of the two halves of the centre tapped secondary. Further D is the duty cycle w.r.t period T and not w.r.t to period T_S. Always D and T should correspond to the waveform at AC (*i.e.* V_{AC}).

APPENDIX - V

The reader can note that solution of magnetic components design is not unique, as it depends on the choice of many parameters like J, K_w, B_m, η, ΔI in inductor etc.. Unless otherwise stated, take J = 3 A/mm^2, B_m = 0.2 T, K_w = 0.4, D_{max} = 0.45, V_D = 1.5 V, η = 0.8, ΔI = 10%.

Design Exercises

1. Design a transformer for the forward converter configuration as shown in fig. A5.1 for the following specifications.

   Input to main rectifier, capacitor filter network = 170V to 270V rms 50 Hz

   Output V_{01} = 5V

   Output current I_{01} = 2A

   Output voltage V_{02} = 12V

   Output current I_{02} = 100 mA

   Switching frequency = 40 KHz

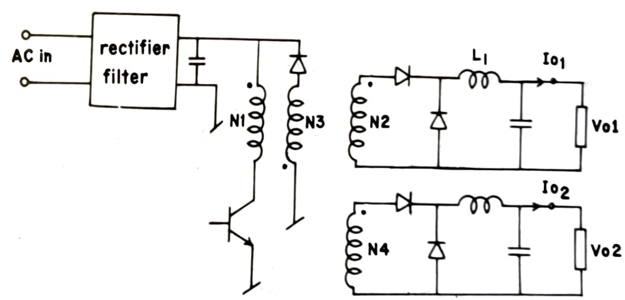

Fig. A5.1