4 PULSE-TRANSFORMER DESIGN AND FABRICATION

Pulse transformers find wide use in a number of electronic circuits. In high-power transmitters they are often used to couple the output of a line-type modulator to the load, and to a lesser extent high-power transformers are utilized in hard-tube modulators. These high-power pulse transformers are typically step-up transformers, which provide a voltage and impedance transformation from their primary to the secondary.

The design of pulse transformers for high-power modulators may be approached in a number of different ways. However, all of the design procedures and approaches are designed to produce a transformer that satisfies a number of differing, and often contradictory, criteria [2,4–20,22–25]. These criteria include the following:

- Achieving specified rise time
- Providing adequate pulse flatness
- Providing desired fall time and tail-of-pulse response
- Establishing conditions for proper tube operation
- Providing voltage transformation

- Holding temperature within reasonable limits
- Providing for a path of filament current for the tube
- Withstanding the required operating voltages
- Small size, weight, and cost
- Withstanding the required voltage stresses

The simultaneous optimization of all these differing requirements may not be possible, and the skillful tradeoff among these parameters is an important part of the design process.

4–1 EQUIVALENT CIRCUIT AND ANALYSIS OF BEHAVIOR

Pulse transformers come in a wide variety of winding configurations, core materials, and turns ratios. A "complete" equivalent circuit that is generally applicable is shown in Figure 4–1 [8,11,23]. As can

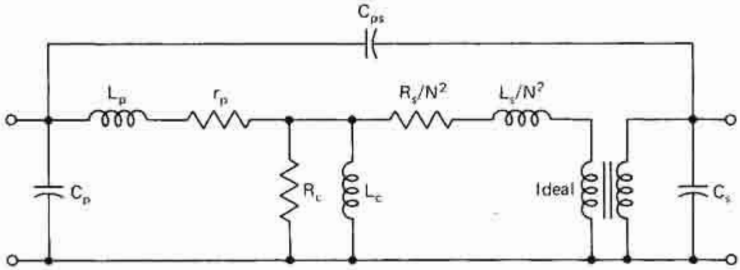


FIGURE 4–1 "Complete" equivalent circuit for a pulse transformer.

be seen, while the analysis of this circuit is not prohibitively complex, the specific pulse transformer performance is highly dependent upon a number of different ratios of the various equivalent circuit elements. In many cases, a substantial simplification is made possible by assuming that the turns ratio of the transformer is greater than 1:3. For such a condition, a considerably simplified equivalent circuit may be utilized, as shown in Figure 4–2 [6,7], where, for simplicity, all circuit elements have been transformed to the same side of the transformer, either the primary or the secondary.

Analysis of this equivalent transformer may be simplified by dividing the pulse into three distinct parts: the rise time, the top of the pulse, and the fall time, or tail of the pulse. For the analysis of rise time, the magnetizing current may usually be neglected; the result is the simplified equivalent circuit shown in Figure 4–3. It may normally be assumed that a step voltage is impressed on the primary of the pulse transformer that gives an output rise time related to the leakage inductance and the stray capacitance of the transformer, the distributed capacity of the

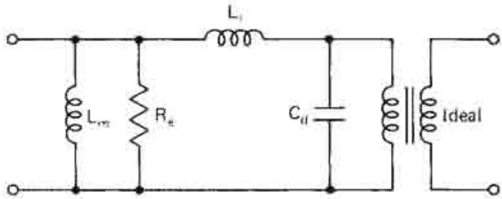


FIGURE 4–2 Simplified equivalent circuit for a step-up transformer.

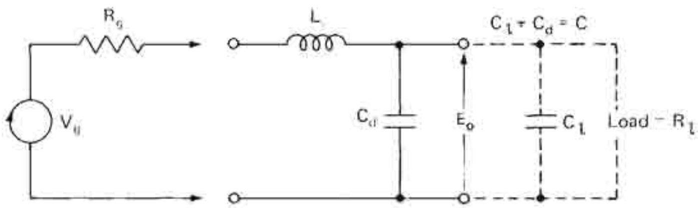


FIGURE 4–3 Simplified equivalent circuit for determining rise time of a pulse transformer.

load, and the nature of the load, i.e., whether it is a resistance load or a biased-diode load. Analysis of the circuit for the underdamped case shows that the output voltage is given by

$$E_o(t) = \frac{V_g R_l}{R_g + R_l}\left[1 - e^{-at}\left(\frac{a}{\omega}\sin \omega t + \cos \omega t\right)\right]$$

where $2a = \dfrac{R_g}{L_l} + \dfrac{1}{CR_l}$

$$b = \frac{1}{L_l C}\left(1 + \frac{R_g}{R_l}\right)$$

the damping coefficient k_1 is given by

$$k_1 = \frac{a}{\sqrt{b}}$$

and the frequency

$$\omega = \sqrt{b - a^2}$$

Figure 4–4 shows that a good compromise between fast rate of rise and minimum overshoot is obtained by a choice of the damping coefficient k_1 between 0.5 and 0.75. When operating into a biased-diode load it is often desirable to choose

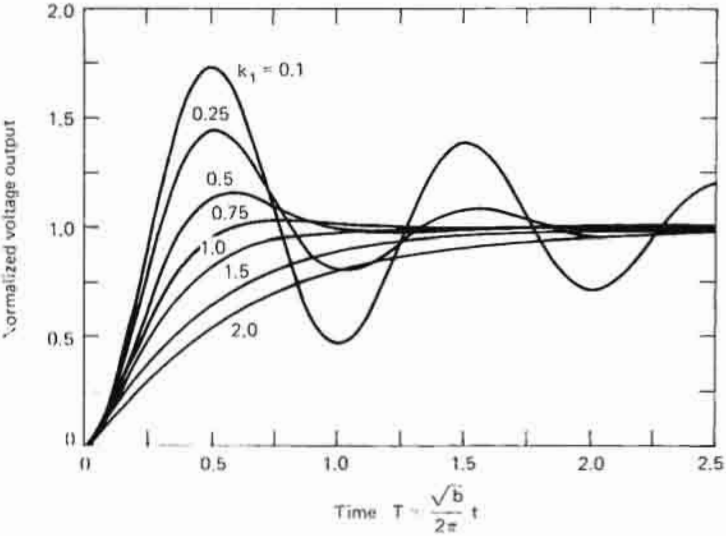


FIGURE 4–4 Output voltage of transformer given by circuit of Figure 4–3 for a range of values of damping coefficient k_1. See text for details of axis labeling.

$$R_l = \sqrt{\frac{L_l}{C_d + C_l}} \tag{4–1}$$

so that current flowing in the leakage inductance just before the biased diode conducts is equal to the current through the load after it conducts. If the conditions of Equation 4–1 are met with a hard-tube modulator (where usually $R_g \ll R_l$ and $CR_g \ll L_l/R_l$), k_1 will equal 0.5; for the case of the line type modulator where $R_g = R_l$, k_1 will equal 0.71. Thus, the usual design procedure is to design for k_1 between 0.5 and 0.75. If k_1 is approximately equal to 0.5, then for a resistive load, the rise time will be given by

$$t_r = 1.78\sqrt{L_l(C_d + C_l)}$$

If the load is a biased diode ($R_l \gg R_g$ during the rise time), the rise time for such a biased diode load is given by

$$t_r = 1.3\sqrt{L_l(C_d + C_l)}$$

Analysis of the performance on the top of the pulse assumes that the voltage is approximately constant and the value of droop in tube current is related to the magnetizing current in the self-inductance of the transformer. For a biased-diode load, the decrease in device current during the pulse is approximately equal to the magnetizing current, i.e.,

$$I_m \approx \Delta I_d$$

where I_m = magnetizing current
ΔI_d = decrease in device current

while the droop in voltage for a linear resistance load is approximately

$$\frac{\Delta E}{E_o} = \frac{I_m}{2I_o}$$

Analysis of performance for the tail of the pulse is somewhat more complex. For proper operation of the transmitter, it is desired that the voltage fall rapidly to zero, have no spurious oscillations, have a low value of backswing, and never assume the same polarity as the initial output voltage. Analysis of the tail-of-pulse performance is complicated by the fact that both the magnetron and the transformer magnetic core material exhibit distinctly nonlinear properties during this time. This behavior may result in a distinctly underdamped condition for the pulse transformer during recovery, possibly giving rise to oscillations on the voltage pulse that can produce RF pulses other than and in addition to the main transmitted pulse. Figure 4–5 shows a sketch of such a

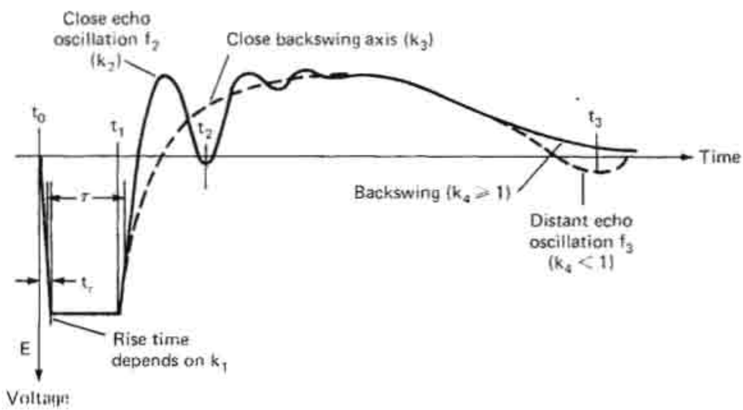


FIGURE 4–5 Line-type modulator output waveform showing oscillations and zero recrossings. [16]

voltage waveform, showing undesired oscillations both close to and distant from the main pulse.

The nonlinear behavior of the magnetron has been discussed earlier; the tube displays distinctly different properties depending upon the voltage applied to the tube. The *BH* curve for core materials used in the pulse transformer during the pulse, as shown in Figure 4–6, shows that

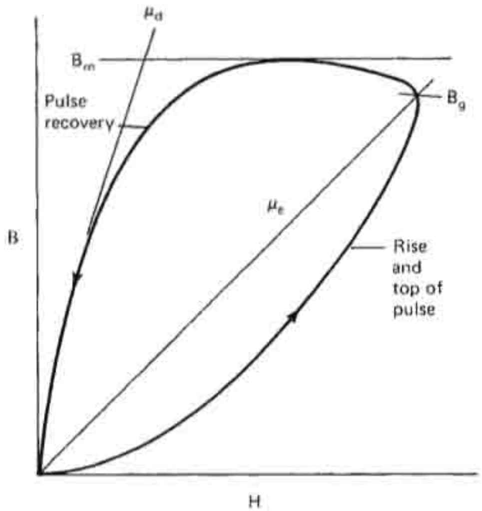


FIGURE 4–6 Pulse permeability of the output transformer core throughout the output pulse. [16]

permeability μ_e at the end of the pulse may be distinctly different from the permeability that occurs during the trailing edge of the pulse, designated μ_d.

Lee [16] has applied approximation techniques for analysis of this particular condition, and utilizes the equivalent circuit shown in Figure 4–7 for the analysis. In this equivalent circuit, C_n is the total pulse-

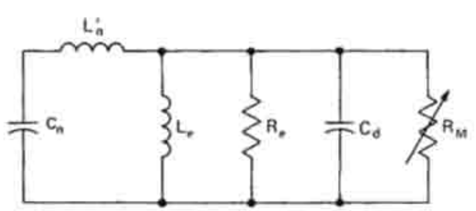


FIGURE 4–7 Equivalent circuit for line-type modulator used for analysis of tail-of-pulse response. [16]

forming network capacitance, L_n' is the sum of the leakage inductance and the PFN inductance, C_d is the stray capacitance of the transformer and load, R_m is the magnetron static resistance, and R_e is the equivalent resistance of the transformer. In addition, a factor J, the ratio of μ_e to μ_d, is also defined, along with an equivalent load resistance during the time interval close to the pulse, R_m. These parameters are then used to calculate three normalized constants, which can be used to check for the presence of close echoes (backswings across the axis close to the pulse), or distant false echoes or axis recrossings. These parameters are summarized in Table 4–1. The value of k_1 was chosen in an earlier

TABLE 4–1 KEY PARAMETERS IN TAIL-OF-PULSE ANALYSIS [16]

Part of pulse affected	Value of load resistance	Impedance ratio defined*	Condition for good pulse shape
Front edge	R_m	$k_1 = \frac{1}{2R_m}\sqrt{\frac{L_l}{C_d}}$	$k_1 = 0.5$ for minimum t_r with flat-top current pulse
Close echo	$R_i = \sqrt{R_m R_e}$	$k_2 = \frac{1}{2R_i}\sqrt{\frac{L_n'}{C_d'}}$	$k_2 \geq$ value in Fig. 4–8 for no close echo
Backswing axis close to pulse	$R_i = \sqrt{R_m R_e}$	$k_3 = \frac{1}{2R_i}\sqrt{\frac{L_e}{C_d'}}$	$k_3 > k_4$ by definition
Distant echo	R_e	$k_4 = \frac{1}{2R_e}\sqrt{\frac{L_e J}{C_d}}$	$k_4 \geq 1$ for no distant echo

* C_d' is the parallel combination of C_d and C_n, all referred to the same side of the transformer.

portion of the design procedure in order to achieve acceptable front-edge performance. In order to check for the presence of close echoes, k_2 and k_3 are calculated. A value of k_3 and the ratio Δ of the magnetizing current to load current are entered into Figure 4–8, and the values of k_2 must be greater than the value obtained from Figure 4–8 for no close echo to exist. Finally, k_4 should be greater than or equal to 1 for no distant echoes to be present. Those interested in details of the derivation are referred to Lee [16] for a more complete discussion.

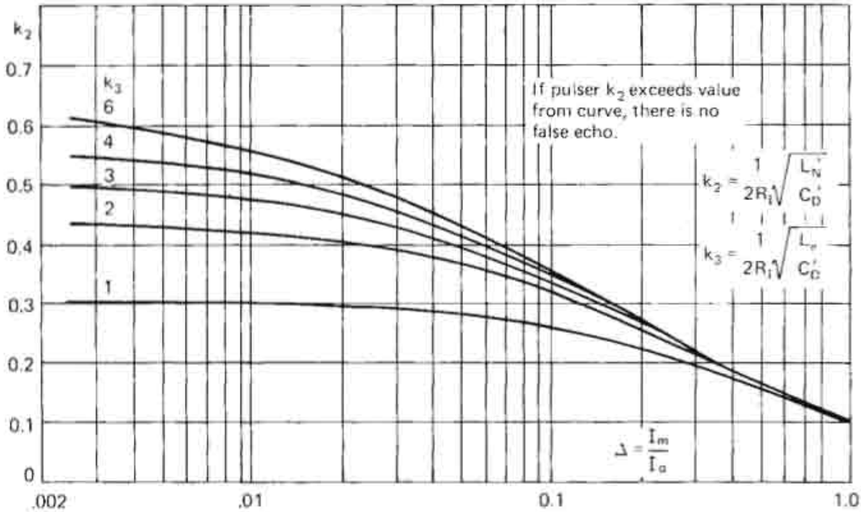
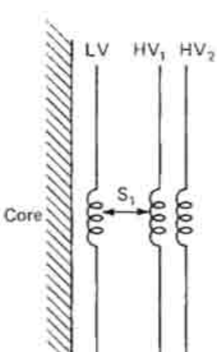


FIGURE 4–8 Graph for determining border line of close false echoes. [16]

4–2 WINDING CONFIGURATIONS

There are a number of possible transformer configurations, several of which are summarized in Figures 4–9 through 4–14, along with formulas for leakage inductance and stray capacitance for each configuration. It should be remembered that the leakage inductance and stray capacitance may be referred to either the low-voltage or the high-voltage side of the transformer. Values may be transformed to the other side of the transformer by dividing all inductances and multiplying all capacitances by the square of the transformer turns ratio. All values in Figures

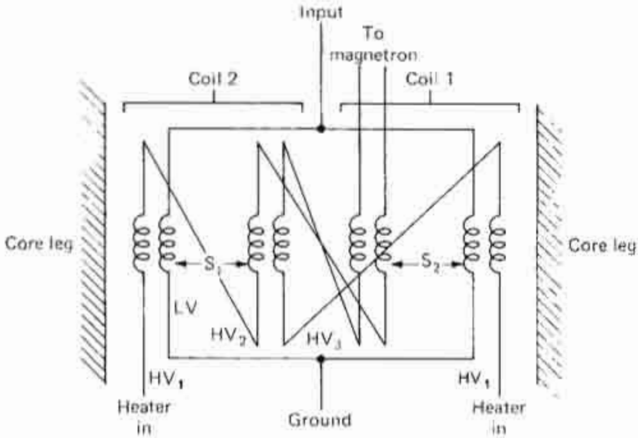


$$L_l = \frac{0.032 N_s^2 l_c}{t}\left(S_1 + \frac{\Sigma d}{3}\right)$$

$$C_d = \frac{0.225 k l_c t}{S_1}\left(\frac{1}{3} - \frac{2}{3n} + \frac{1}{3n^2}\right)$$

1. Traverse for all windings the same.
2. Layers HV_1 and HV_2 have same number of turns.

FIGURE 4–9 Single-layer-primary, single-layer-secondary transformer winding. [6]

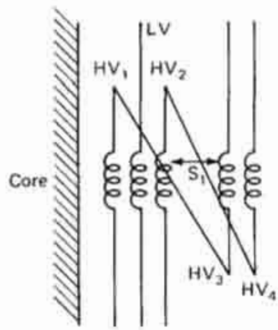


$$L_l = \frac{0.032 N_s^2 l_c}{t} \times 1/4 \times \left(\frac{n-1}{n}\right)^2 \times \left(S_1 + S_2 + \frac{\Sigma d}{3}\right)$$

$$C_d = 0.225 k l_c t \left(\frac{1/12 + 1/4n^2}{S_2} + \frac{7/12 - 1/2n + 1/4n^2}{S_1}\right)$$

1. All traverse the same.
2. HV_1 and LV have same number of turns.
3. HV_2 and HV_3 have same number of turns.
4. HV_1 and LV are two wires wound side by side.

FIGURE 4–10 Lord-type transformer winding. [6]



$$L_l = \frac{0.032 N_s^2 l_c}{t}\left(\frac{n-1}{n}\right)^2\left(S_1 + \frac{\Sigma d}{3}\right)$$

$$C_d = \frac{0.225 k l_c t}{S_1}\left(\frac{1}{3} - \frac{1}{3n} + \frac{1}{3n^2}\right)$$

1. All traverse the same.
2. HV_1, LV, HV_2 have same number of turns.
3. HV_3, HV_4 have same number of turns.

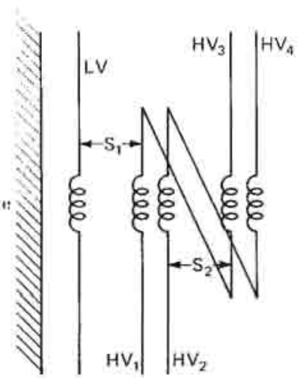
FIGURE 4–11 Interleaved-primary–single-secondary transformer. [6]

4–9 through 4–14 are referred to the high-voltage winding. The definitions of the symbols in Figures 4–9 through 4–14 are

C_d = distributed capacity of high-voltage winding (pF)
L_l = leakage inductances referred to the high–voltage winding (μH)

N_s = total number of high-voltage winding turns series-connected
l_c = average mean length of turn (in)
t = wire traverse (in)
n = ratio of high-voltage winding turns to low-voltage winding turns
S = insulation pad thickness (in)
k = relative dielectric constant of insulation
d = radial build of the copper of a winding layer, when the winding layer carries pulse current (in)

A representative cross section of a pulse transformer with single-layer primary and bifilar-wound single secondary is given in Figure 4–15, which shows the principal portions of the transformer and the appropri-

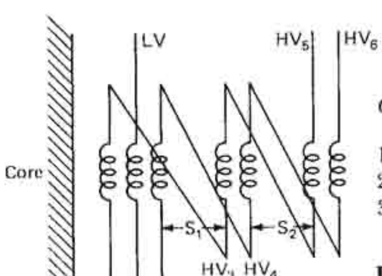


$$L_l = \frac{0.032 N_s^2 l_c}{t}\left(S_1 + \frac{S_2}{4} + \frac{\Sigma d}{3}\right)$$

$$C_d = 0.225 k l_c t\left(\frac{1/12 - 1/3n + 1/3n^2}{S_1} + \frac{1/4}{S_2}\right)$$

1. All traverse the same.
2. HV_1, HV_2, HV_3, and HV_4 have same number of turns.

FIGURE 4–12 Isolated-two-layer-secondary transformer. [6]

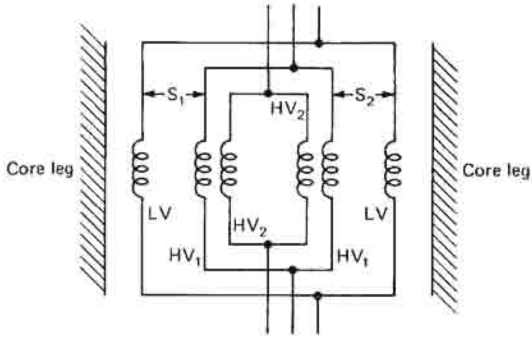


$$L_l = \frac{0.032 N_s^2 l_c}{t}\left(\frac{n-1}{n}\right)^2\left(S_1 + \frac{S_2}{4} + \frac{\Sigma d}{3}\right)$$

$$C_d = 0.225 k l_c t\left(\frac{1/12 + 1/4n^2}{S_1} + \frac{1/4 - 1/2n + 1/4n^2}{S_2}\right)$$

1. Traverse same for all windings.
2. HV_1, LV, and HV_2 have same number of turns.
3. HV_3, HV_4, HV_5, and HV_6 have same number of turns.

FIGURE 4–13 Interleaved-primary, double-layer-secondary transformer. [6]



$$L_l = \frac{0.032 N_s^2 l_c}{t} \times \frac{1}{4}\left(S_1 + S_2 + \frac{\Sigma d}{3}\right)$$

$$C_d = 0.225 k l_c t \left(\frac{1/3 - 2/3n + 1/3n^2}{S_2} + \frac{1/3 - 2/3n + 1/3n^2}{S_2}\right)$$

1. All traverse the same.
2. $S_1 = S_2$.
3. HV_1 and HV_2 have same number of turns.

FIGURE 4–14 Paralleled primary and secondary transformer winding. [6]

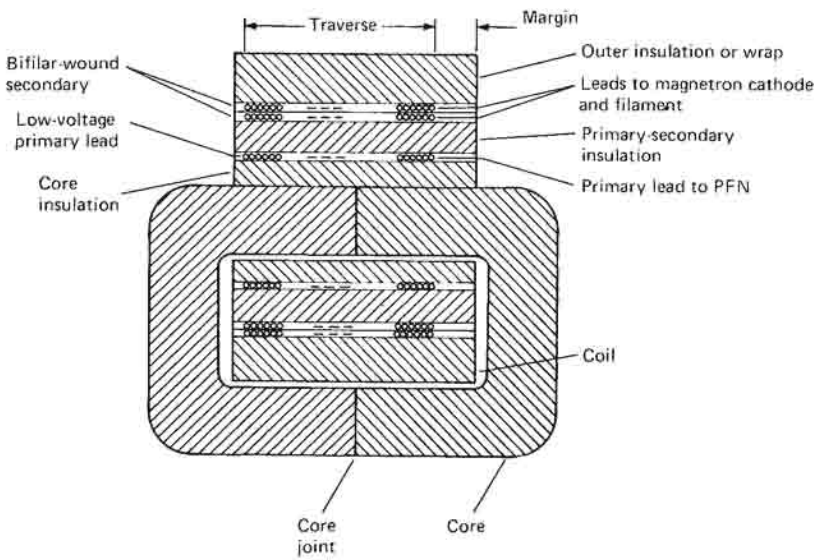


FIGURE 4–15 Cross-sectional view of a typical simple (single-layer primary, single-layer, bifilar-wound secondary) high-voltage pulse transformer, showing major features and nomenclature.

ate terminology. A large portion of the art of pulse-transformer design involves the choice of winding configuration, and dimensional ratios that result in desired values of L_l and C_d [5–7,9–11,14,15,17–20,24,25]. The detailed design of a high-power pulse transformer, however, involves a number of different parameters, including heat transfer, and physical winding configuration, all of which are treated in the following sections.

4–3 PROPERTIES OF MAGNETIC CORE MATERIALS

Almost all high-power pulse transformers are wound on cores of magnetic material. The most common configuration is the so-called C core, which is shown schematically in Figure 4–16. This core is made of grain-oriented silicon steel strips (Silectron), a thickness of 2 mils being a common value for use in high-power pulse transformers. Such material is available commercially under a number of different names, including Silectron, Hypersil, Magnesil, Microsil, and Supersil. A summary table of dimensions of a number of typical pulse-transformer C cores is included in Table 4–2.

The pulse permeability of the transformer core is a function of pulse length and flux change in the core. A summary of pulse permeability from Fenoglio et al. [6] is given in Figure 4–17, a manufacturer's typical data are given in Figure 4–18 [2], and the core loss per pulse is summarized in Figure 4–19.

There are cases when optimum permeability may be obtained only by shimming the air gap in the core or by applying a resetting magnetomotive force (mmf). This condition usually occurs only for those conditions for which the core is large or the required flux changes are large. The introduction of an air gap reduces the residual induction of the core, resulting in a larger ΔB for a given value of pulse magnetizing force H_m. If a resetting force is available, it may not be necessary to gap the core. By applying resetting mmf, it is possible to bias the core to achieve a large value of ΔB.

In order to obtain data such as are shown in Figures 4–17 and 4–18, the number of volts per turn is calculated for a given value of ΔB by the equation

$$V/N = \frac{6.45AS\,\Delta B}{t \times 10^8}$$

where V = peak voltage at end of pulse (V)
N = number of turns
A = gross core area in^2 ($D \times E$)
S = stacking factor (0.89 for 2-mil Silectron)

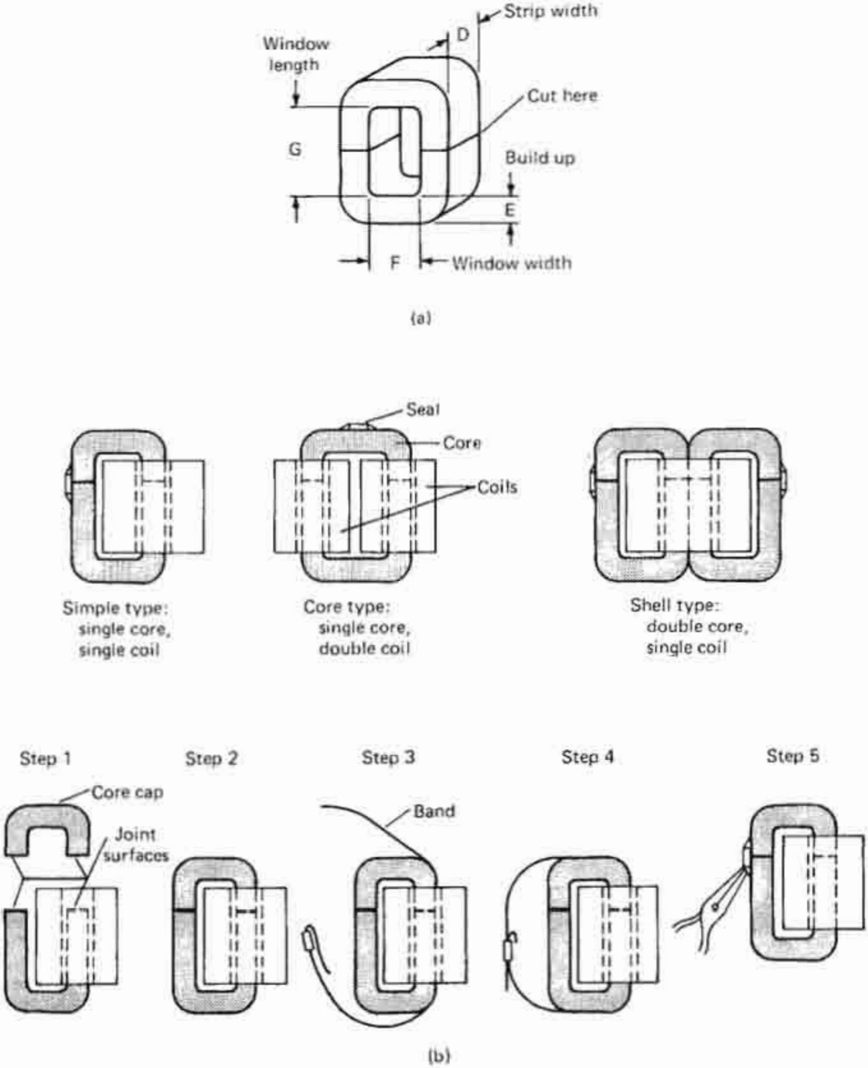


FIGURE 4–16 Cut C-core (*a*) nomenclature and (*b*) banding data. [2]

t = pulse length (s)
ΔB = induction change in gauss (G)

The pulse permeability is then calculated as

$$\mu_e = \frac{\Delta B \times l \times 2.54}{0.4\pi N I_m}$$

TABLE 4–2 REPRESENTATIVE SET OF 2-MIL C CORES [2]

	Dimensions (in or in²)					
Part no.	Strip width *D*	Build *E*	Window width *F*	Window length *G*	Core area $D \times E$	Core length $2F + 2G + 2.9E$
L-6	1/2	1/4	1/4	7/8	0.125	2.98
L-8	3/8	3/8	3/8	1 3/16	0.141	4.21
L-9	1/2	3/8	3/8	1 3/16	0.188	4.21
L-13	5/8	1/4	1/2	1 1/8	0.156	3.98
L-10	5/8	3/8	3/8	1 3/16	0.234	4.22
L-12	1/2	7/16	1/2	1 1/8	0.219	4.40
L-11	3/4	3/8	3/8	1 3/16	0.281	5.62
L-78	3/4	5/16	5/16	2 1/4	0.234	6.03
L-14	1/2	1/2	1/2	1 9/16	0.250	5.58
L-18	1/2	7/16	5/8	1 9/16	0.219	15.64
L-15	5/8	1/2	1/2	1 9/16	0.313	5.58
L-16	3/4	1/2	1/2	1 9/16	0.375	5.58
L-17	1	1/2	1/2	1 9/16	0.500	5.58
L-19	1	1/2	5/8	1 9/16	0.500	5.83
L-20	1	5/8	5/8	1 9/16	0.625	6.19
L-24	1	5/8	3/4	2 5/16	0.625	7.94
L-25	1	7/8	15/16	2 1/2	0.875	9.41
L-248	1 1/8	3/4	1 1/8	2 7/8	0.844	10.18
L-98	1	5/8	2	3	0.625	11.81
L-54	2	3/4	3/4	4	1.5	11.68
AL-1079	4	4	8 1/2	16	16	60.60

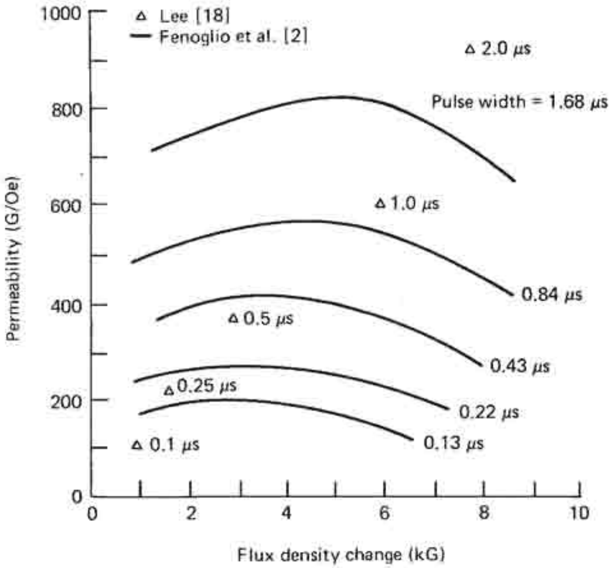


FIGURE 4–17 Pulse permeability vs. flux-density change for 2-mil Silectron.

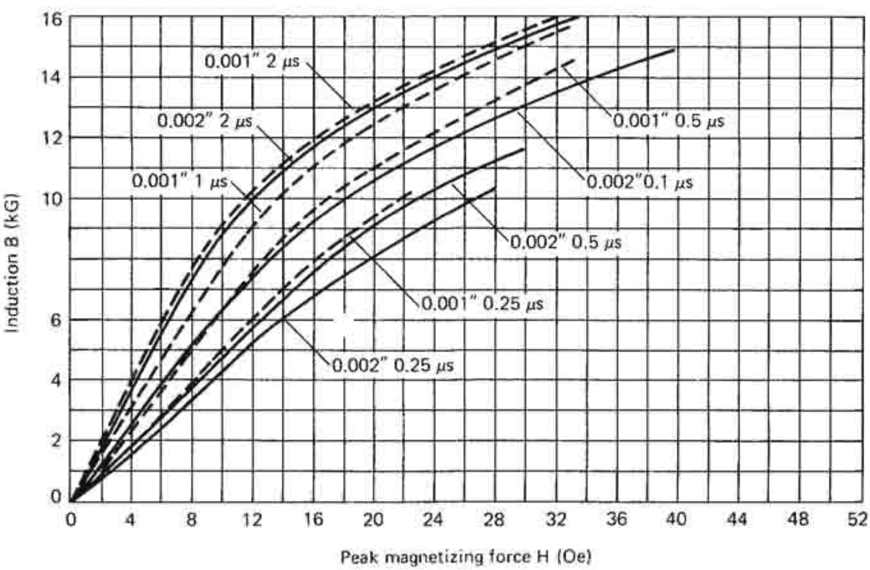


FIGURE 4–18 Typical pulse magnetization curve for Silectron. [2]

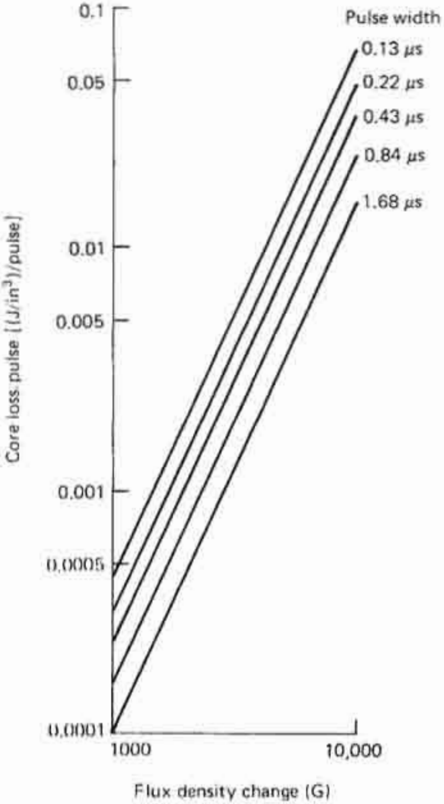


FIGURE 4–19 Pulsed core losses for 2-mil Silectron [6]. To obtain core loss in watts, multiply the number of joules per cubic inch per pulse by core volume (in³) by the number of pulses per second.

where μ_e = effective pulse permeability in gauss per oersted (G/Oe)
I_m = peak exciting current (A)
l = core length (in) $(2F + 2G + 2.9E)$

4–4 INSULATION

While dry types of solid insulation may be used below the 10- to 15-kV range, above that level vacuum-impregnated liquid-solid-composite insulation is used almost exclusively. The most common liquids are the transformer oils and the silicone fluids, and commonly used solids are kraft paper, Mylar, and Teflon. Properties of commonly used insulation are summarized in Table 4–3. In some cases, the insulation thickness and margins should be selected on the basis of physical strength rather than dielectric strength; Table 4–4 tabulates such thicknesses.

TABLE 4–3 PROPERTIES OF COMMONLY USED INSULATION MATERIALS

Material	Dielectric constant	Maximum working stress: Puncture	Maximum working stress: Creep (along surface)
Kraft paper in oil	4.5	50 V rms/mil low-frequency AC; 250–300 V/mil pulsed for less than 50 kV applied; 200–225 V/mil over 50 kV pulsed	20–30 V/mil pulsed less than 50 kV and low-frequency AC; 10 V/mil over 50 kV pulsed
Mylar in oil	3.4	400 V/mil pulsed	30 V/mil pulsed
Teflon in oil	2.3	300–500 V/mil pulsed	30 V/mil pulsed

In no event should the insulation consist of less than two layers of insulating material. The core tube is normally selected for strength and layered with insulating material. Because of irregularities and sharp edges, core tubes are often stressed at one-half to one-third the normal insulation stress.

TABLE 4–4 MINIMUM LAYER INSULATIONS AND MARGINS FOR PHYSICAL STRENGTH

Wire size	Layer insulation (in)	Margins (in)
11–14	0.010	7/32
15–18	0.007	13/64
19–22	0.005	3/16
23–26	0.004	11/64
27–30	0.003	5/32
31–34	0.002	9/64

4–5 HEAT TRANSFER, COOLING, AND THERMAL DESIGN

There are several methods of calculating the hot-spot temperature in transformers [6,11]. Some rather exact methods are available, but these are somewhat involved. If one desires only an approximate answer, some simplifying assumptions may often be made. In the following procedure, simplifying assumptions have been made, and the hot-spot temperatures calculated will be conservative. If more exact values are desired, the methods given in the references are appropriate.

Calculation of Temperature Rise

The first step in calculating the temperature rise in a transformer is to make a careful sketch of the problem and label all pertinent temperature rises, as shown in Figure 4–20. We now assume that the losses at various places in the transformer have been calculated.

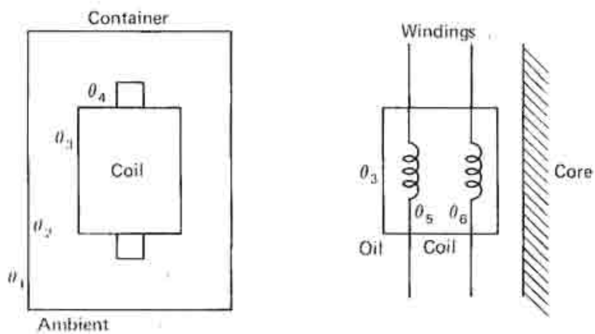


FIGURE 4–20 Simplified schematic representation of temperature drops in an enclosed transformer.

1. The drop from case to ambient θ_1 is calculated as follows. Calculate the outside-envelope area of the can. This is determined by multiplying the string distance around the can by the can height and adding the product to the cover area. The base area is not included unless it is known to be in direct contact with a suitable cold surface. The total heat produced inside the can is divided by the envelope area to find a heat-flux density. An equivalent temperature rise is then determined by a combined radiation and convection curve such as the one in Figure 4–21.

2. θ_2, the drop from case to oil, is calculated as follows. Calculate the inside tank area that is contacted by the oil. Determine

the heat-flux density H_c for this area by dividing the total number of watts by this area. From a suitable curve such as Figure 4–22, determine θ_2.

3. θ_3, the drop from coil to oil, is calculated by using the same methods as were used to find θ_2. The area used is only the *vertical* surface of the coil.

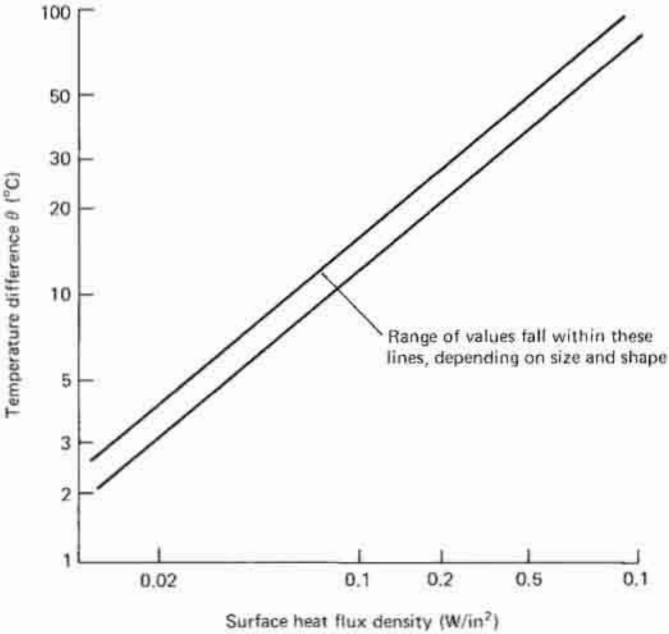


FIGURE 4–21 Total heat transfer by both convection and radiation at atmospheric pressure.

4. θ_4, the drop from core to oil, is computed by assuming the heat leaves the edges of the core laminations and by using the methods used to find θ_2 and θ_3.

5. θ_5 and θ_6 are conduction rises and are calculated by the equation

$$\theta = W r_t \lambda$$

where θ = temperature difference (°C)
W = rate of heat flow (W/in²)
r_t = thermal resistivity of insulation, (°C/in)/(W/in²)
λ = length of heat-flow path (in)

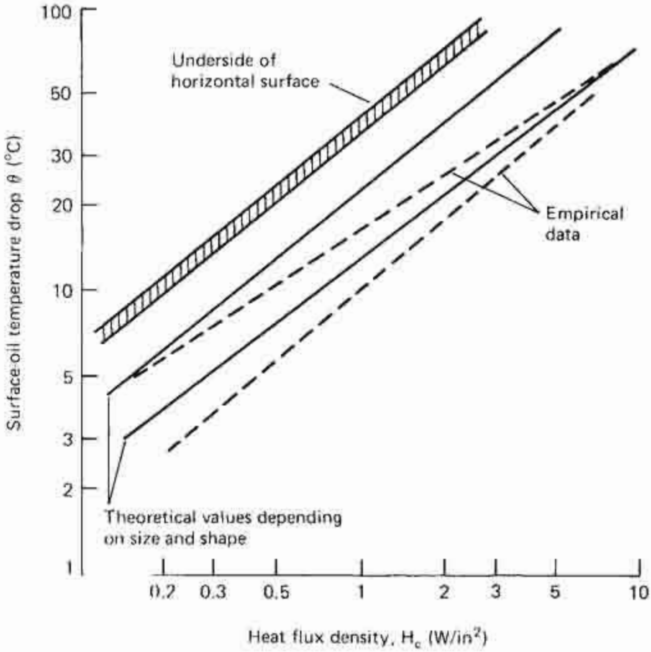


FIGURE 4-22 Natural convection heat transfer in transformer oil.

It is usually assumed that all heat flows outward in the coil. Typical values for r_t are:

Material	r_t, (°C/in)/(W/in²)
Kraft paper, oil-impregnated	250
Water	70
Air	1710
Transformer oil (noncirculating)	245
Mica	110
Teflon	170
Nomex (nylon paper)	160–230

The sum of all rises from any point in the coil to the ambient must not exceed the allowable temperature rise for the insulation system being used. For kraft paper the maximum temperature is normally taken to be 105°C for 10,000 h life or 130°C for 2000 h life; oil temperature should be limited to 100°C maximum.

In determining temperature rise, consideration should be given to the expansion of oil with temperature. The coefficient of expansion for oil is 0.073% °C^{-1}; for a 100°C rise, there will be over a 7% increase in oil volume. For small units with moderate temperature rises, the expansion may be accommodated by deformation of the can. Otherwise, the can surfaces may be especially "dimpled" to permit expansion, gas-expansion space may be provided, or an expansion bellows may be incorporated. Sometimes, glass balls or nylon chips (resin) may be added to reduce oil volume; care should be taken that this filler material does not adversely affect the heat-transfer or voltage-breakdown properties of the unit.

Sources of Heat

There are several sources of heat in a pulse transformer, principally:

- Core losses
- Energy stored in stray capacitances
- Losses in conductors

Core losses are calculated by reference to Figure 4–19, and by multiplying the value obtained by the core volume and the PRF to obtain the power dissipated in the core. It is normally assumed that the heat is transferred to the oil only by the exposed core surface.

Energy stored in stray capacitances may contribute significantly to transformer loss. Energy stored is given by the quantity

$$\frac{1}{2} CV^2$$

which when multiplied by the PRF gives the total power. If the load is resistive, much of this energy is dissipated in the load; however, if it is a biased-diode load, this is not the case. In a line-type modulator, the clip-diode circuit may absorb much of the energy in the stray capacitances. Each individual case must be considered separately, and between 10% and 90% of this stored energy may be dissipated in the transformer.

Conductor losses are of two types: those due to pulse currents, and those due to filament current. Filament current calculations may use dc resistance values for power-dissipation calculations. For pulse currents, the situation is somewhat more complex.

Size (AWG)	Area (cmil)	Diameter			Turn/in (1 layer enamel)	Ω/1000 ft at 20°C 100% cond.	Lb per 1000 ft, bare
		Bare-wire	1 layer enamel	2 layers enamel or heavy polymer			
10	10,380	0.1019	0.1039	0.1056	9	0.999	31.4
11	8,234	0.0907	0.0927	0.0943	10	1.260	24.9
12	6,530	0.0808	0.0827	0.0842	11	1.588	19.88
13	5,178	0.0720	0.0738	0.0753	12	2.003	15.68
14	4,107	0.0641	0.0659	0.0673	13.5	2.525	12.43
15	3,257	0.0571	0.0588	0.0602	15	3.184	9.86
16	2,583	0.0508	0.0525	0.0539	17	4.016	7.82
17	2,048	0.0453	0.0469	0.0482	19	5.064	6.20
18	1,624	0.0403	0.0418	0.0432	21	6.385	4.92
19	1,288	0.0359	0.0374	0.0387	24	8.051	3.90
20	1,022	0.0320	0.0334	0.0346	27	10.15	3.09
21	810	0.0285	0.0300	0.0310	30	12.80	2.45
22	624.4	0.0263	0.0267	0.0278	34	16.14	1.94
23	509.5	0.0226	0.0238	0.0249	39	20.36	1.54
24	404.0	0.0201	0.0213	0.0224	43	25.67	1.22
25	320.4	0.0179	0.0191	0.0201	48	32.37	0.970
26	254.1	0.0159	0.0170	0.0180	54	40.81	0.769
27	201.5	0.0142	0.0153	0.0161	60	51.47	0.610
28	159.8	0.0126	0.0136	0.0145	67	64.90	0.484
29	126.7	0.0113	0.0122	0.0130	75	81.83	0.384
30	100.5	0.0100	0.0109	0.0116	84	103.2	0.304
31	79.70	0.0089	0.0100	0.0105	94	130.1	0.241
32	63.21	0.0080	0.0088	0.0095	104	164.1	0.1913
33	50.13	0.0071	0.0078	0.0085	117	206.9	0.1517
34	39.75	0.0063	0.0070	0.0075	131	260.9	0.1203

The basic equations for determining the effective pulse resistance of conductors are rather complex and are not particularly complete [6,7,11,15]. In many cases, significant simplifying assumptions can be made. For sinusoidal excitation,

$$R_{eff} = 2.47 \times 10^{-7} \sqrt{f} \times \frac{\text{length of wire}}{\text{bare-wire diameter}}$$

where f is the frequency in Hz. For pulses,

$$R_{eff} = 1.58 \times 10^{-7} \frac{1}{\sqrt{\tau}} \times \frac{\text{length of wire}}{\text{bare-wire diameter}} \tag{4–2}$$

where τ is the pulse length in seconds [10]. These expressions assume that skin effects are important. To check this assumption, assume that skin effects are important and calculate the resistance. Then the dc resistance should be looked up and the larger of the two resistances should be used.

There is a proximity effect that must also be considered [11,10]. If current flows only on one-half of the wire, these values of the resistance must be multiplied by 2. This is the case for the Lord-type transformer-winding configuration. It should be noted that for many pulse transformers, the conductor losses are normally dominated by filament current losses.

4–6 BUSHINGS AND FABRICATION TECHNIQUES

Fabrication Constants

The winding traverse is obtained by using the actual wire width, obtained from wire tables such as Table 4–5, and multiplying by approximately 1.15. Similarly, the radial build is obtained by multiplying the insulation and wire build by 1.25. Specific values for these constants may vary with individual winding technique.

Bushings

Normally, the transformer is hermetically sealed in a container and the leads are brought out through bushings. Some representative high-voltage solder bushings are described in Tables 4–6 and 4–7. In general, the flashover voltage for a bushing is given approximately by $17.5t^{0.63}$kV, where t is the length of the surface creep path in inches; in general, the bushing should be worked at one-third to one-fourth of this stress [21]. If high-altitude operation, or operation under conditions of excessive dust or humidity, is contemplated, voltage ratings

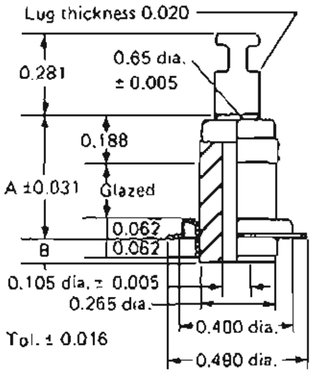
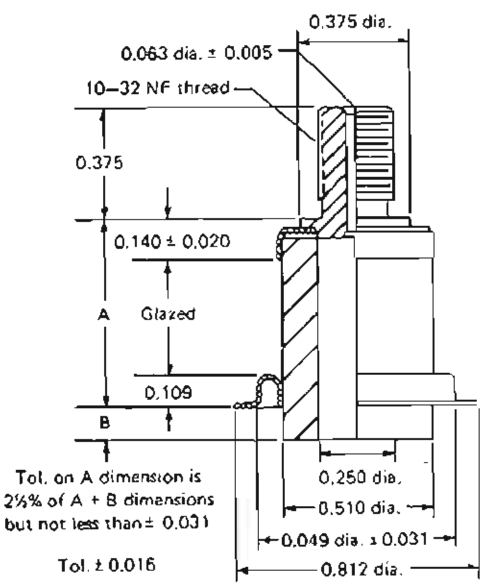


TABLE 4-6 SOME LOWER-VOLTAGE INSULATED BUSHINGS [3]

Dimensions (in)		Average flashover (kV rms)	Average corona start (kV rms)	Approx. net weight (oz)
A	B			
0.460	0.094	6.6	6.5	⅛
0.580	0.094	8.1	6.6	⅛
0.460	0.219	6.6	6.5	⅛
0.814	0.188	11.4	6.7	¼



Dimensions (in)		Average flashover (kV rms)	Average corona start (kV rms)	Approx. net weight (oz)
A	B			
0.458	0.031	6.6	4.0	¼
0.562	0.031	8.4	4.9	¼
0.625	0.094	9.5	5.4	¼
0.656	0.031	10.1	5.6	¼
0.741	0.406	11.4	6.1	¼
0.758	0.031	11.7	6.2	½
0.806	0.141	12.5	6.4	½
0.859	0.031	13.3	6.7	½
0.875	0.250	13.4	6.8	½
0.953	0.031	14.8	7.2	½
0.984	0.188	15.2	7.3	½
1.150	0.203	17.5	8.0	½
1.156	0.031	17.5	8.0	½
1.188	1.125	17.9	8.1	½
1.250	0.031	18.8	8.3	½
1.356	0.031	20.0	8.6	½
1.656	0.125	23.6	9.5	¾
1.688	0.312	23.9	9.6	¾
1.957	0.250	26.5	10.3	¾
2.357	0.250	29.0	11.4	1
2.656	0.344	30.8	11.9	1⅛

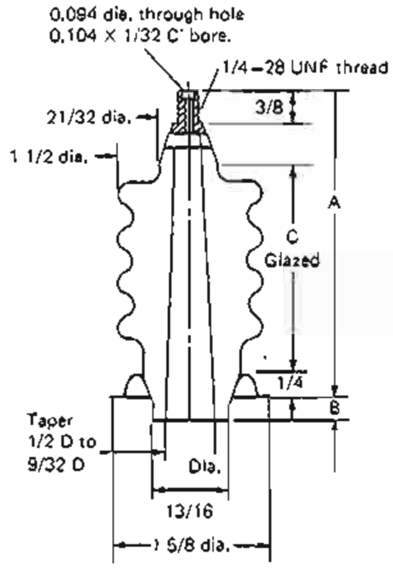
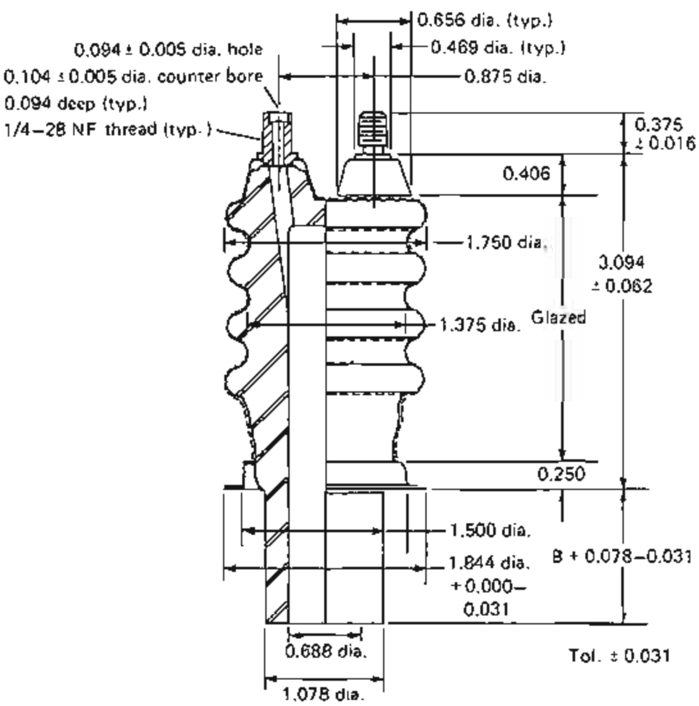


TABLE 4-7 SOME LARGE HIGH-VOLTAGE CERAMIC BUSHINGS [1.3]

No. of flutes	A	B	C	Average flash-over (kV)	Approx. net weight (oz)
2	2 9/32	1/4	1 1/4	25	4
3	2 25/32	1/4	1 3/4	30	5
4	3 9/32	1/4	2 1/4	35	6
5	3 25/32	1/4	2 3/4	40	7 1/2
5	3 25/32	1	2 3/4	40	8



Type	Dimensions (in) B	Average flashover (kV rms)	Average corona start (kV rms)	Approx. net weight (oz)
As shown on drawing	0.500	36.9	16.8	9
With 0.091-in dia. by 5-in long OFHC copper conductors	0.500	36.9	16.8	9 1/2
As shown on drawing	1.250	36.9	16.8	10
With 0.091-in dia. by 5.7-in long OFHC copper conductors	1.250	36.9	16.8	10 1/2

should be reduced accordingly. For some applications, cathode bushings of the microwave tube may be inserted into a special ceramic well, eliminating any exposed output bushings. In such cases, care must be used to ensure that the cathode stem is adequately cooled.

Vacuum Impregnation

In order to achieve satisfactory operation at high voltages, the transformer must be thoroughly dried, completely impregnated, and hermetically sealed in a leakproof container. Considerable care is necessary in order to avoid air pockets, or voids, in the insulation system; the presence of voids results in high electric fields and possible breakdown and deterioration of the insulation system. Therefore, voids are to be avoided if at all possible. A typical impregnation sequence is described in the following paragraphs.

Take the component to be impregnated and place it in an oven at 90°C for 4 h minimum. The component should be completely sealed except for one small fill hole.

The oil should be heated under a vacuum for 3 h at 70°C to prepare it. The hot component should then be pumped to less than 1 mmHg for 2 h at 90°C and the oil then allowed to fill the container while maintaining a vacuum of less than 40 mmHg vacuum; the component should remain under a vacuum for 0.5 h. Then it should sit at atmospheric pressure for 0.5 h. Then the component should undergo four cycles of 15 min at 40 mmHg and 15 min under normal atmospheric pressure (or even positive pressure). During all of these procedures the oil should be maintained at approximately 70°C.

While the component is still under oil and the temperature at 70°C, the fill hole should be solder-sealed. The component should then be removed from the oil and thoroughly cleaned (trichlorethylene is useful here). Place the component in a refrigerator to reduce the temperature to 4°C (or some other appropriate temperature) and then place it in an oven at 90°C until thoroughly heated. The component should then be checked to see if any leaks are apparent (powdered talc makes leaks more evident); if no leaks are visible, the unit may be painted.

These procedures are not to be considered inviolate, and variations from these procedures may be made with no dire consequences. This procedure has been used successfully and is based on empirical data from the power-transformer industry.

4–7 Pulse-Transformer Design Procedures

The design of pulse transformers for use in line-type or hard-tube modulators is rather complex, and no single design procedure is

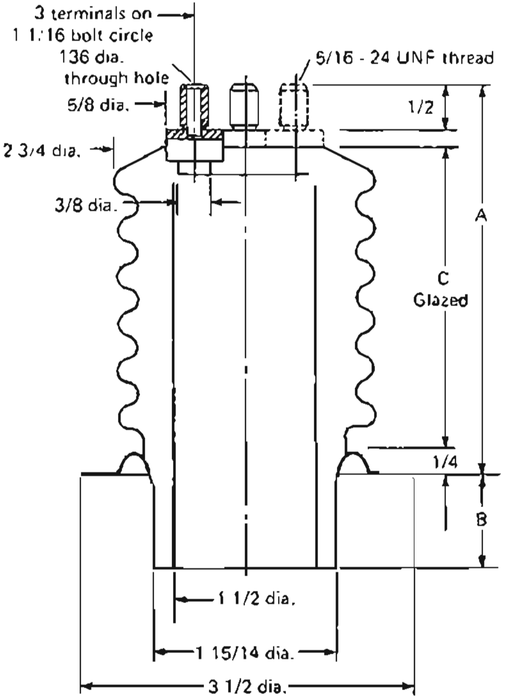


TABLE 4–7 *Continued*

No. of flutes	A	B	C	Average flash-over (kV)	Approx. net weight
2	2 3/16	1/2	1 1/4	26	10 oz
3	2 11/16	1/2	1 3/4	31	13 oz
4	3 3/16	1/2	2 1/4	36	1 lb
5	3 11/16	1/2	2 3/4	41	1 1/4 lb
6	4 3/16	1	3 1/4	45	1 1/2 lb
8	5 3/16	1 1/4	4 1/4	50	2 lb

applicable for all cases. Rather exhaustive studies of the problems involved are available [5–7,10,12,14–20,24,25], and the serious designer should become thoroughly familiar with these works. For many ordinary transformers (pulse length of 0.2 to 2 μs, maximum output voltage less than 25 kV, turns ratio $\approx$ 5, PRF $\leq$ 5 kHz), a rather simple design procedure will sometimes be adequate.

The design of a transformer involves the selection of suitable core, number of turns, winding configuration, wire size, and insulation thickness. There are many sophisticated design procedures available, but they are all merely ways to make a good first choice of core size and number of turns. In our design procedure, we will assume the transformer will be used in a line-type modulator, utilize a trial-and-error method of core selection, and then calculate the number of turns required to achieve a suitable value of exciting current at the end of the pulse (10% is a reasonable value). A winding configuration is assumed, and a coil is designed by using suitable values for creep and puncture stress for the insulation used and a suitable wire size for the currents involved. The dimensions of the coil are then adjusted to try to achieve "optimum" values of leakage inductance and distributed capacity. The coil is then checked to see that the tail-of-pulse behavior is satisfactory and that the hot-spot temperature is not excessive. If all of these criteria are met, the transformer design is then considered complete. The actual test of any transformer design is how well it operates in the circuit for which it is intended, and the actual operation of the transformer must be considered an integral part of any design procedure.

First, select a core from a list of available 2-mil silicon steel cores (Hypersil, Selectron) [2]. Then assume an initial value of flux density change (using Figure 4–23) and the corresponding permeability (Figures 4–18 and 4–19). The number of turns, the flux density, and the exciting current are related by the equation

$$\frac{V}{N}=\frac{6.45AS\,\Delta B}{t\times 10^{8}} \tag{4–3}$$

where V = peak voltage of end of pulse (V)
N = number of turns
A = gross core area (in^2)
S = stacking factor (0.89 for 2-mil Silectron)
t = pulse length (s)
ΔB = induction change (G)

and

$$I_m=\frac{2.02\,\Delta B\,l}{N\mu_r} \tag{4–4}$$

where μ_e = effective pulse permeability
I_m = peak exciting current (A)
l = core length (in)

The exciting current should never exceed 10% of the peak pulse current at the end of the pulse.

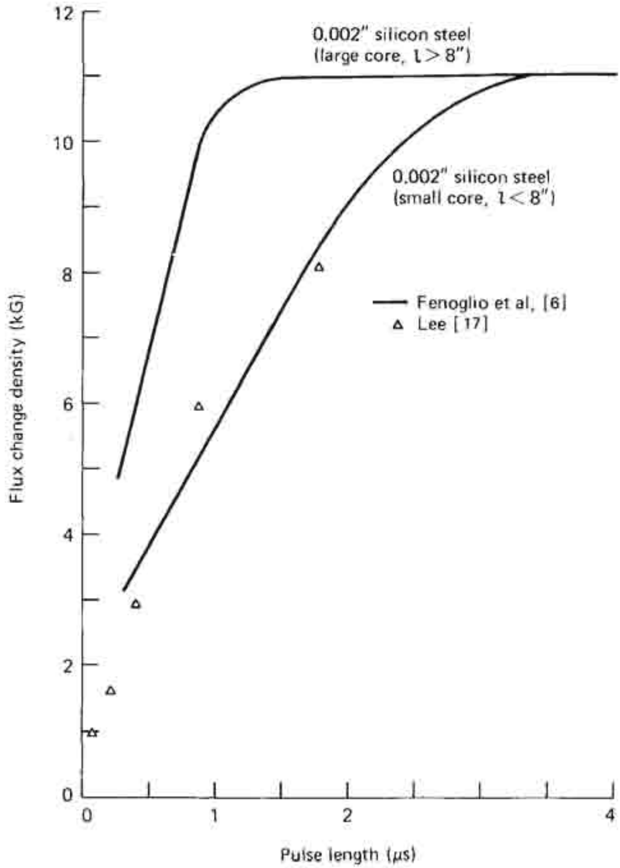


FIGURE 4–23 Suggested initial flux change as a function of pulse length.

The winding configuration most useful for our purposes might be one shown in Figure 4–9 or 4–10. The initial choice of wire size is usually made on the basis of the amount of window available for winding by using a 90% space factor. Experience has shown that if the resulting current density in the wire exceeds 1 A/1500 cmil, the hot-spot temperature may be excessive. The radial build of the finished coil may be obtained by multiplying the sum of the insulation and wire thickness in the radial direction by about 1.25. Paralleling smaller wire sizes should not be overlooked as a favorable alternative to extremely large wire sizes, and the wire sizes should be adjusted until all windings have the same length. The initial choice of insulation thickness should be made on the basis of the allowable stress in the dielectric. Suitable interlayer-insulation stresses under pulsed service are as follows [10]:

Insulation	Puncture stress (V/mil)	Creep stress (V/mil)	Dielectric constant
Kraft paper in oil	200–300	20–30	4.5
Mylar in oil	400	30	3.4
Teflon in oil	300–500	30	2.3

A reasonable stress for core tube insulation is 25 V/mil.

The leakage inductance and distributed capacity of the coil should now be calculated. For the Lord-type winding (Figure 4–10),

$$L_l = \frac{0.032 N_s^2 l_c}{4t}\left(\frac{n-1}{n}\right)^2 \times \left(S_1 + S_2 + \frac{\Sigma d}{3}\right)$$

$$C_d = 0.225 k l_c t\left(\frac{1/12 + 1/4n^2}{S_2} + \frac{7/12 - 1/2n + 1/4n^2}{S_1}\right)$$

where C_d = distributed capacity of high-voltage winding (pF)
L_l = total leakage inductance referred to the high-voltage winding (μH)
N_s = total number of high-voltage winding turns in series connection
l_c = average mean length of turn (in)
t = wire traverse (in)
n = ratio of high-voltage winding turns to low-voltage winding turns
S_1, S_2 = insulation pad thickness (in) (see Figure 4–10)
d = radial build of the copper of a winding layer, when the winding layer carries pulse current (in)
k = dielectric constant of coil insulation

The exact values of leakage inductance and distributed capacity desired are not obvious, but if one specifies for a magnetron load that

$$t_r = 1.3 \times 10^{-3} \sqrt{L_l\,(C_d + C_l)}$$

where t_r = rise time of transformer when source has zero rise time (μs)
C_d = distributed capacity of the transformer (pF)
L_l = leakage inductance of the transformer (μH)
C_l = capacity of the load, including bushing and wiring capacity (pF)

and that

$$R_p = 10^3 \sqrt{\frac{L_l}{C_d + C_l}}$$

where R_p = static resistance of the magnetron (kΩ)

the simultaneous solution of these equations gives

$$L_l = \frac{t_r}{1.3} \times R_p \tag{4–5}$$

$$C_d = \frac{t_r \times 10^6}{1.3 \times R_p} - C_l \tag{4–6}$$

The dimensions of the coil should be adjusted within the limitations of wire size, core size, and properties of the dielectric in an attempt to achieve these values.

The tail-of-pulse response may be checked by the method of Lee [16], which was outlined earlier. The only nonobvious parameters are μ_e, μ_d, and R_e. R_e is usually calculated from the values of core loss (Figure 4–20), and μ_e and μ_d may be obtained from Figures 4–18 and 4–19.

The temperature rise of the coil should now be checked to ensure that it is not excessive, by using procedures described earlier in this section. In making these calculations, one should remember that the pulse current flows on the inner half of the wires and the secondary wire connected to the magnetron cathode carries essentially all of the magnetron current pulse [6]. The values of core loss under pulsed conditions may be found from Figure 4–19, or may be calculated as $(2/3)E_o I_m \times$ duty cycle, in watts.

A final part of the design procedure must include actual operation in the circuit to verify compliance with the design objectives. If any of the objectives are not met, the transformer must be redesigned with either a different core, a different number of turns, a different winding configuration, or all three.

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