

Optimal Design of Square Wave Pulse Transformer with Short Rise Time

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DOI: <https://doi.org/10.51583/IJLTEMAS.2025.1409000088>

Abstract: Pulses used in radar systems, linear accelerators, and in Klystron and magnetrons require square waves with short rise times and small overshoot.

However, the high-voltage pulse transformer designed and manufactured has a large leakage inductance and a constant distributed capacitance, which results in a long rise time and a large overshoot.

In this paper, an analysis of the structure size of the pulse transformer to reduce the rise time of the pulse and propose a method to obtain the optimum result were investigated.

And the correctness of the proposed design results is verified by analyzing the relationship between the influence of core material and switching speed and the Klystron (electron tube) with nonlinear inductance characteristics.

Keywords: Pulse transformer, Rise time, Optimal design, Genetic algorithm

I. Introduction

High-voltage pulse technology is widely used in waste water treatment, synthetic fuel production processes, sterilization, radar systems and linear accelerators, and its applications are expanding day by day.

And the design and operation of high-voltage pulse transformers is a prerequisite for the development of the machine building industry and the electronics industry.

Most devices to generate pulse use IGBT modules, which are applied in series or in parallel with output devices with high switching speed in case of high output power and high voltage.

In case of output devices to be connected in parallel, it is important to be able to provide protection and control about overcurrent and overvoltage, and optimal design of pulse transformer.

In document [1], a pulse modulator generating a pulse in the μs -range and a typical pulse waveform (Figure 1) are shown.

If the current in a parallel-connected power device is balanced, the duration of the pulse is minimized.

In addition, the transformation ratio of pulse transformer makes to select the rated voltage and current of the switching device in the design of the pulse modulator conveniently.

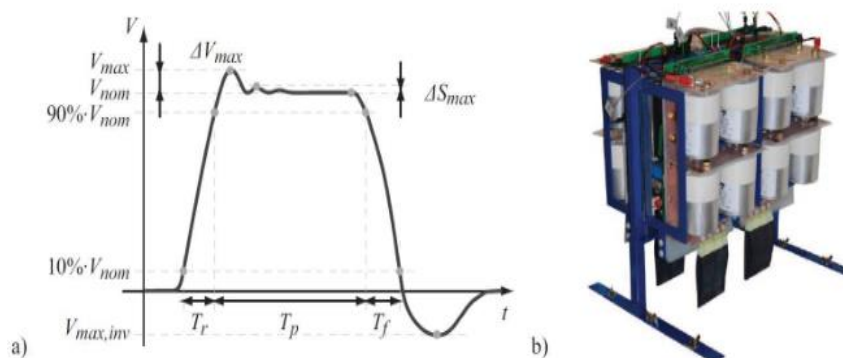


Figure 1. Pulse modulator and output pulse waveform.

(a) Output pulse waveform; (b) 20MW pulse modulator

On the other hand, the parasitic capacitances of the pulse transformer generate transient voltages, affect the rise time and the magnitude of the overshoot ΔV_{max} , which complicates the design of the transformer.

Since the permeability of the magnetic material is not ideal as $\mu \neq \infty$, the maximum flux density B_{max} is small and the minimum critical frequency f_u of the pulse transformer is determined, corresponding to the maximum voltage generation time.

The leakage inductance of the pulse transformer depends on the magnetic coupling between the primary and secondary windings and, in combination with the parasitic capacitance, produces f_0 above the critical frequency.

In general, parasitic capacitances are synthesized with a single capacitance C_d .

The output voltage close to the square wave looks like a broad frequency spectrum.

In order to transmit a voltage impedance with low pulse distortion and maximum width at the rising edge, the parasitic capacitance of the pulse transformer should be reduced.

Finally, the pulse transformer is an important component of the pulse modulator that affects the rise time T_r of the output pulse and the overshoot ΔV_{max} .

This paper analyzes the influence of leakage inductance L_σ and C_d , which main parameters of the pulse transformer, and shows the process of calculating the rise time and overshoot of the pulse transformer, and propose an optimal design method to reduce rise time of pulse transformer.

And through the design and test analysis of the pulse transformer used for plasma ignition were verified the accuracy of the optimal design method.

II. Equivalent circuit of pulse transformer

Figure 2-a is shown the normalized equivalent circuits of low and high-frequency pulse transformers introduced in many literatures.

To simplify the analysis of transient processes operating at square wave voltages, the normalized equivalent circuit is simplified as shown in Figure 2-b and the transformation ratio is $n \gg 1$.

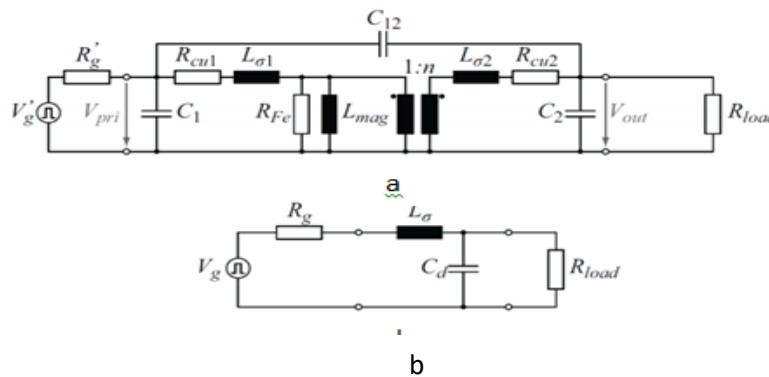


Figure 2. Equivalent circuit of pulse transformer. Normalized equivalent circuit (a) Simple equivalent circuit. (b)

In the equivalent circuit, all the inductive resistance and input voltage U_g are calculated to the secondary and the symbol '»' is attached.

In case of rise time of the pulse $T_r \approx 100\text{ns}$, the voltage generation time is very short, so that the core is not excited during that period.

Thus, the effect of R_{Fe} and L_{mag} on the rise time is negligible.

The resistance component of the core- R_{Fe} does not affect the rise time of the pulse.

For example, in 20 MW pulse transformer, since the resistance of the core is connected in parallel with the load resistance, it is $R_{load} = 1500\Omega$ if referred to the secondary and $R_{load,pri} = 0.052\Omega$ if referred to the primary.

There, the load resistance R_{load} which was calculated to the primary is much smaller than the resistance of core $R_{Fe} \approx 4\Omega$.

Therefore, the rise time and the overshoot of output voltage are mainly determined by the leakage inductance L_σ and capacitance C_d .

If an ideal square-wave voltage is added to the primary winding, the output voltage $U_{out}(t)$ is as follows. [Document 4]

$$U_{out}(t) = \frac{U_g R_{load}}{R_g + R_{load}} \left[1 - e^{-\sigma t} \left(\frac{a}{k} \sinh(kt) + \cosh(kt) \right) \right] \quad (1)$$

Here

$$k^2 = a^2 - b$$

$$2a = \frac{R_g}{L_\sigma} + \frac{1}{C_d R_{load}}, \quad b = \frac{1}{L_\sigma C_d} \left(1 + \frac{R_g}{R_{load}} \right)$$

The damping coefficient σ in Equation (1) is

$$\sigma = \frac{a}{\sqrt{b}} = \frac{C_d R_g R_{load} + L_\sigma}{2\sqrt{R_{load} L_\sigma C_d (R_g + R_{load})}} \quad (2)$$

If the waveform of the output pulse is basically determined by the transformer characteristics, the impedance R_g of the modulator can be neglected.

Thus, the damping coefficient σ which only considers the effect of the transformer can be simplified as follows :

$$\sigma = \frac{a}{\sqrt{b}} = \frac{1}{2R_{load}} \sqrt{\frac{L_\sigma}{C_d}} \quad (3)$$

As in the load parasitic capacitance C_{load} , the parasitic inductance of the pulse generator L_{gen} and the parasitic capacitance C_{gen} must consider certainly for the calculation of the rise time and the overshoot which depend on the transformation ratio of the pulse transformer and appear in the connection line between the DC bus and the switching devices. Then the input impedance is changed from the resistance R_g to the impedance Z_g .

If the transformation ratio of the pulse transformer is large, the parasitic capacitance C_{gen} is neglected, and if it is calculated to a secondary, the capacitance is reduced to $1/n^2$.

Thus, the parasitic capacitance of the transformer C_d and the capacitance of the load C_{load} are smaller.

Figure 3 is shown the transient response of the normalized output voltage during $T = \frac{\sqrt{b}}{2\pi} t$.

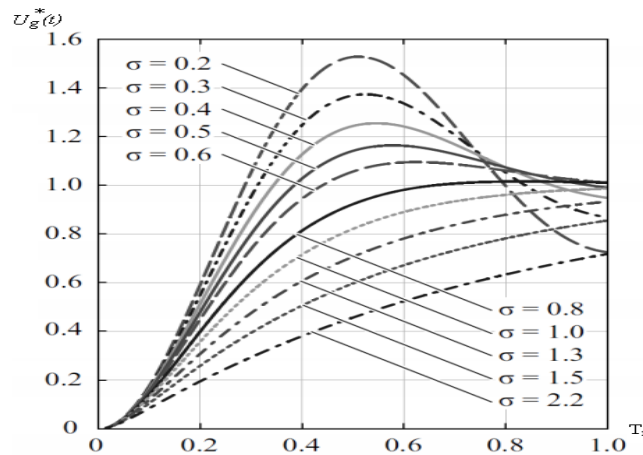


Figure 3. Transient response of output voltage versus damping coefficient.

As shown in Figure 3, the relationship between T_r and overshoot begins to appear clearly since $\sigma < 1$.

If the damping coefficient σ is decreased, the rise time T_r will be shorter.

Therefore, to minimize the rise time T_r , the maximum value of the overshoot should be within the allowable range ($< 3\%$) and makes damping coefficient as minimum as possible.

Overshoot

From Equation (3), the damping coefficient σ depends on L_σ and C_d and depends on the structural dimensions of the pulse transformer and load impedance R_{load} .

In general, the load impedance R_{load} changes according to the application object.

Therefore, the structural dimensions of the pulse transformer must design to satisfy the characteristics of the pulse waveform.

For example, in Klystron with a load of $R_{load} = 1500\Omega$, the allowable maximum overshoot is 3% and a damping coefficient $\sigma = 0.75$ is necessary.

Finally, the R_{load} and σ are given, the ratio of leakage inductance L_σ and capacitance C_d is determined as follows :

$$2R_{load} \cdot \sigma = \sqrt{\frac{L_\sigma}{C_d}} \quad (4)$$

Rise time

In addition to the overshoot, the rise time of output voltage T_r can be obtained from Equation (1).

$$T = \frac{\sqrt{b}}{2\pi} t, \quad T_r = 2\pi T_{10\%-90\%} \sqrt{L_\sigma C_d} \quad (5)$$

As shown in Equation (5), T_r is proportional to L_σ and C_d .

The coefficient $T_{10\%-90\%}$ depends on the chosen damping coefficient σ and equals the time when the voltage $U_{load(t)}$ rises from 10% to 90% as shown in Figure 3.

For $\sigma=0.75$, the coefficient is $T_{10\%-90\%}=0.365$.

Since the rise time T_r is proportional to $L_\sigma \cdot C_d$, the parasitic parameters must be small for the rise time as short as possible.

For example, in order to keep the rise time below $T_r=500\text{ns}$, $L_\sigma \cdot C_d$ should be less than 4.75×10^{-14} in case of $\sigma=0.75$.

If the load impedance is $R_{load}=1\ 500\Omega$ and the ratio of L_σ to C_d is fixed, the maximum value in Klystron is $L_\sigma < 490\mu\text{H}$ and $C_d < 97\text{pF}$ separately.

In order to satisfy the requirement for maximum transient deviation and minimum rise time, L_σ and C_d must have maximum values at a given ratio of L_σ to C_d .

In general, the pulse modulator is connected to the primary winding, as is the case when the load connected to the secondary winding is $\frac{L_{gen}}{C_{load}}$.

Indeed, the parasitic inductance of the pulse generator was measured as $L_{gen}=260\mu\text{H}$.

The typical capacitance values of Klystron is in range of $C_{load}=40\sim 120\text{pF}$.

This means that the leakage inductance and distributed capacitance of the pulse transformer should be small to satisfy the pulse characteristics.

Thus, Equations (4) and (5) were expressed as follows.

$$\begin{cases} 2R_{load} \cdot \sigma = \sqrt{\frac{L_\sigma + L_{gen}}{C_d + C_{load}}} \\ T_r = 2\pi T_{10\%-90\%} \sqrt{(L_\sigma + L_{gen})(C_d + C_{load})} \end{cases} \quad (6)$$

III. Optimal design of pulse transformer

Design characteristics of pulse transformer

The ratio of L_σ to C_d can vary depending on the structural dimensions of the pulse transformer.

It depends on the insulation distance, width, height, and length of the winding.

However, the magnitude of L_σ and C_d is determined by the geometrical form of the pulse transformer and it can consider constant.

Therefore, in first the geometrical form of the pulse transformer should be chosen so that the magnitude of $L_\sigma \cdot C_d$ is reduced.

Then, the structure dimensions should be designed so that the required ratio of L_σ to C_d is obtained.

In the following, the magnitude of $L_\sigma \cdot C_d$ for the three transformer of geometrical form is analyzed.

The leakage inductance L_σ and capacitance C_d are calculated from the magnetic and electric field energies.

$$E_{mag} = \frac{1}{2} \mu \int_V \vec{H}^2 dV \equiv \frac{1}{2} L_\sigma I_{Pri}^2 \quad (7)$$

$$E_{elec} = \frac{1}{2} \epsilon \int_V \vec{E}^2 dV \equiv \frac{1}{2} C_d U_{Pri}^2 \quad (8)$$

To simplify the comparison, only the energy between the windings is considered.

Figure 4 is shown the geometrical dimensions of the pulse transformer for the three winding types.

1) Cylindrical winding

Cylindrical windings are used widely due to their simple structure.

The primary and secondary windings are wound on two insulating cylinders and are separated as the required insulation distance.

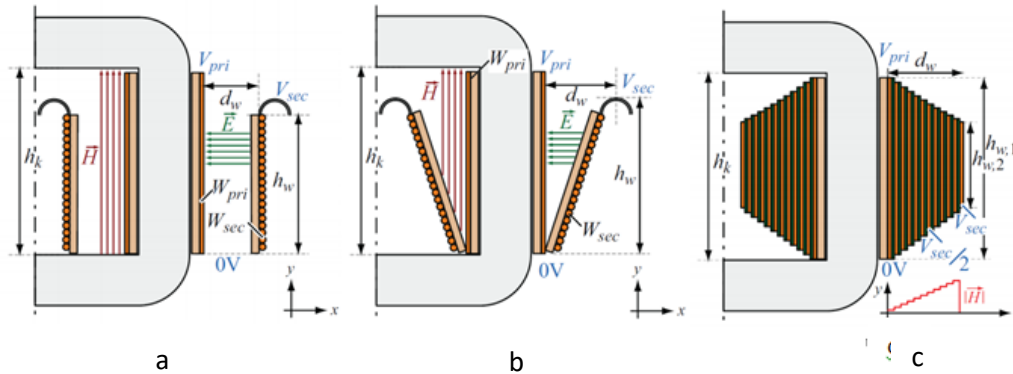


Fig. 4. Geometrical dimensions of the pulse transformer according to the type of winding

a) cylindrical winding b) conical winding c) laminar winding

The leakage inductance is basically determined by the magnetic field strength and volume between the insulating cylinders from Equation (7).

According to Ampere's law and assuming the permeability $\mu = \infty$ in an ideal core material, the magnetic field strength H in the core window is calculated by the product of the current of the primary winding and the number of turns $N_{pri} \cdot I_{pri}$ and the height h_k of the core window.

$$\vec{H} = \frac{N_{pri} I_{pri}}{h_k} \quad (9)$$

Using Equations (7) and (9), the magnetic field energy E_{mag} stored between the primary windings W_{pri} and W_{sec} is calculated as follows.

$$E_{mag} = \frac{1}{2} \mu (N_{pri} I_{pri})^2 \frac{l_w \cdot d_w}{h_k} = \frac{1}{2} L_{\sigma} I_{Pri}^2 \quad (10)$$

Finally, the leakage inductance of the cylindrical winding is

$$L_{\sigma, parallel} = \mu \frac{N_{pri}^2 l_w \cdot d_w}{h_k} \quad (11)$$

To calculate the capacitance C_d , the linear voltage distribution $U_{pri}(y)$ and $U_{sec}(y)$ is assumed across the winding.

$$U_{pri}(y) = \frac{y}{h_w} U_{pri}; \quad U_{sec}(y) = \frac{y}{h_w} U_{sec} \quad (12)$$

In addition, the voltage difference between the primary and secondary winding depends on the vertical position y .

$$\Delta U(y) = U_{sen}(y) - U_{pri}(y)$$

Due to the voltage difference between the primary and secondary winding, the electric force lines become horizontal as shown in Figure 4-a.

, the electric field energy $E(y)$ depends on the position of y .

$$|\vec{E}(y)| = \frac{\Delta U(y)}{d_w} = \frac{U_{pri}(n-1) \cdot y}{h_w d_w} \approx \frac{U_{sec} \cdot y}{h_w d_w} \quad (13)$$

Considering Equation (8), the electric field energy stored between the primary and secondary winding and the capacitance C_d are calculated as follows :

$$\begin{aligned} E_{elek} &= \frac{1}{2} \epsilon \int_0^{l_w} \int_0^{h_w} \int_0^{d_w} \left(\frac{U_{sec} \cdot y}{h_w d_w} \right)^2 dx dy dz \\ &= \frac{1}{6} \epsilon U_{sec}^2 \cdot \left(\frac{l_w \cdot h_w}{d_w} \right) = \frac{1}{2} C_d U_{pri}^2 \end{aligned} \quad (14)$$

$$C_{d,parallel} = \frac{1}{3} \epsilon \cdot \left(\frac{N_{sec}}{N_{pri}} \right)^2 \cdot \left(\frac{l_w \cdot h_w}{d_w} \right) \quad (15)$$

Thus, the magnitude of $L_\sigma \cdot C_d$ in a pulse transformer with a cylindrical winding is as follows.

$$L_{\sigma,parallel} \cdot C_{d,parallel} = \frac{1}{3} \cdot \epsilon \cdot \mu \frac{N_{sec}^2 \cdot l_w^2 \cdot h_w}{h_k} \quad (16)$$

2) Conical winding

In a transformer with a cylindrical winding the distance between the windings is constant, but the voltage increases linearly in the y-direction, so the electric field between the windings increases linearly.

In order to keep the electric field energy $\vec{E}(y)$ constant, the distance d_w between the windings is linearly reduced so that the voltage difference becomes smaller. To do so, it is reasonable to use the conical winding shown in Figure 4-b.

Compared to the cylindrical winding, the volume between the windings and the leakage inductance L_σ can be reduced, but C_d increases due to the reduced distance between the windings.

According to the finite element analysis, it can be assumed that the magnetic field is constant in the y-direction when calculate the leakage inductance L_σ of the conical winding for $d_w \ll h_w$.

Considering Equation (7), the stored magnetic field energy E_{mag} and the leakage inductance $L_{\sigma,cone}$ are as follows :

$$E_{mag} = \frac{1}{4} \mu (N_{pri} I_{pri})^2 \frac{l_w \cdot d_w}{h_k} = \frac{1}{2} L_\sigma I_{pri}^2 \quad (17)$$

$$L_{\sigma,cone} = \frac{1}{2} \cdot \mu \frac{N_{pri}^2 \cdot l_w \cdot d_w}{h_k} \quad (18)$$

Because the distance $d_w(y)$ and the voltage distribution in the y-direction $\Delta U(y)$ increase linearly, the electric field $\vec{E}$ between the windings is constant and almost parallel to the x-direction as shown in Figure 4-b.

Thus, from Equation (8), the electric energy and capacitance C_d stored in the conical winding are

$$E_{elek} = \frac{1}{4} \epsilon U_{sec}^2 \cdot \left(\frac{l_w \cdot h_w}{d_w} \right) = \frac{1}{2} C_d U_{pri}^2 \quad (19)$$

$$C_{d,cone} = \frac{1}{2} \epsilon \cdot \left(\frac{N_{sec}}{N_{pri}} \right)^2 \cdot \left(\frac{l_w \cdot h_w}{d_w} \right) \quad (20)$$

Finally, the magnitude of $L_\sigma \cdot C_d$ in the conical winding is

$$L_{\sigma, cone} \cdot C_{d, cone} = \frac{1}{4} \cdot \epsilon \cdot \mu \frac{N_{sec}^2 \cdot l_w^2 \cdot h_w}{h_k} \quad (21)$$

Compared to the cylindrical winding, the size of $L_{\sigma} \cdot C_d$ is reduced by 25% and the rise time is improved by 13.4%.

3) Lamellar winding

The geometrical dimension of the pulse transformer with laminated winding is shown in Figure 4-c.

The secondary winding is wound directly on the primary winding and the turns insulating material is a material with very low permittivity.

The insulation thickness d_{iso} is chosen small because the voltage difference between the two turns is $U_{w,w} = U_{sec}/N_{sec}$.

However, the height of the winding decreases linearly from $h_{w,1}$ to $h_{w,2}$ due to the increase in the voltage difference between turns and center.

The total thickness of the winding dw is equal by the thickness of the isolation d_{iso} and the foil d_{cu} times the number of turns.

The leakage inductance L_{σ} of the sheet winding is calculated again from Equation (7) by the stored magnetic energy.

According to Ampere's law, the magnetic field increases gradually with the number of turns n_L because it is proportional to the current placed in the closed circuit (Figure 4-c).

$$\vec{H}(n_L) = \frac{n_L \cdot I_{sec}}{h_k} \quad (22)$$

The total magnetic energy is the sum of all energies between windings and expressed as follows.

$$\begin{aligned} E_{mag} &= \frac{1}{2} \mu U(n_L) \sum_{n_L=1}^{N_{sec}} H(n_L)^2 = \frac{1}{2} \mu \frac{I_{sec}^2}{h_k^2} U(n_L) \sum_{n_L=1}^{N_{sec}} n_L^2 \\ &\approx \frac{1}{4} \mu (N_{pri} I_{pri})^2 \frac{l_w \cdot d_w}{h_k} = \frac{1}{2} L_{\sigma} I_{pri}^2 \end{aligned} \quad (23)$$

Thus, $L_{\sigma, foil}$ is

$$L_{\sigma, foil} = \frac{1}{2} \cdot \mu \frac{N_{sec}^2 \cdot l_w \cdot d_w}{h_k} \quad (24)$$

The capacitance C_d can be calculated by considering the series connection of the plate capacitors $C_{w,w}$ between turns.

The distance between the plates is equal to d_{iso} and is expressed by the thickness of the entire winding as follows :

$$d_{iso} = \frac{d_w}{(k+1) \cdot N_{sec}} \quad (25)$$

$$\text{Here } k = \frac{d_{cu}}{d_{iso}}$$

Assuming the constant winding height $h_w = (h_{w,1} + h_{w,2})/2$, the capacitance C_d of the sheet winding is

$$C_{d, foil} = (k+1) \cdot \epsilon \cdot \left(\frac{N_{sec}}{N_{pri}} \right)^2 \cdot \left(\frac{l_w \cdot h_w}{d_w} \right) \quad (26)$$

Thus, the leakage inductance and the capacitance product $L_{\sigma} \cdot C_d$ are

$$L_{\sigma, foil} \cdot C_{d, foil} = \frac{(k+1)}{2} \cdot \epsilon \cdot \mu \frac{N_{sec}^2 \cdot l_w^2 \cdot h_w}{h_k} \quad (27)$$

Considering only the magnetic and electrical energy stored between the primary winding W_{pri} and the secondary winding W_{sec} , the pulse transformer with conical winding is suitable because the rise time T_r is shortest when $L_{\sigma} \cdot C_d$ is the smallest.

Because the considered volume affect mainly the influence on the magnitude of magnetic and electrical energy, the magnitude of $L_{\sigma} \cdot C_d$ is an important indicator of the pulse transformer design.

Optimal design method of pulse transformer by genetic algorithm

1) Formulation for Optimal Design Problem of Pulse Transformers

(1) Building the objective function

The main goal in the design of the pulse transformer is to achieve efficiency maximization and $L_{\sigma} \cdot C_d$ minimization.

Therefore, the objective function for the optimal design of the pulse transformer should be a multiobjective function that reflects these indices and the individual objective function is chosen as follows :

- Maximizing efficiency

To maximize the efficiency is the main objective of the optimal design.

The efficiency in the transformer is calculated as follows :

$$\eta = \left(1 - \frac{\sum p}{S + \sum p} \right) \cdot 100, \%$$

Here $\sum p$ - total losses in the transformer, S-transformer capacitance.

The total loss in the pulse transformer is calculated as follows :

$$\sum p = P_{\text{copper}} + P_{\text{iron}}$$

Here, P_{Cu} - Copper loss, P_{Fe} - iron loss.

Hence, the objective function for efficiency maximization is expressed as follows.

$$\left. \begin{aligned} \max : \eta &= f_1(x_1, \dots, x_i, \dots, x_m) \\ x_i &\geq 0, m = 3 \\ x_i &\in [x_{\min}, x_{\max}] \end{aligned} \right\} \quad (28)$$

The current density, flux density and shape coefficient are chosen as the design variables.

Then $x_1 = \Delta$ (current density), $x_2 = B_{\text{iron}}$ (flux density), $x_3 = \beta$ (shape coefficient).

- $L_{\sigma} \cdot C_d$ minimization

From the analysis above, it is problem to optimize the parameters of Equation (21), because the magnitude of $L_{\sigma} \cdot C_d$ is the smallest for the pulse transformer with conical winding.

To minimize $L_{\sigma} \cdot C_d$, the product of l_w and h_w should be chosen to minimize, when the secondary turns N_{sec} is given.

However, N_{sec} depends on the flux density, so that it is constrained by the condition for efficiency maximization.

Also, a winding length and height have a close relationship with the shape coefficient β .

Hence, the objective function for $L_{\sigma} \cdot C_d$ minimization is expressed as follows.

$$\left. \begin{aligned} \min : C &= f_2(x_1, \dots, x_i, \dots, x_m) \\ x_i &\geq 0, m = 3 \\ x_i &\in [x_{\min}, x_{\max}] \end{aligned} \right\} \quad (29)$$

Here, the design variables are as follows:

$x_1 = N_{\text{sec}}$ (number of secondary turns)-Multi-objective optimization functions are constructed as individual objective functions over, $x_2 = l_w$ (average length of secondary winding), and $x_3 = h_w$ (height of secondary winding).

In this paper, a unified objective function of a multi-objective optimization problem is constructed based on the multiplicative method.

$$\min : f(X) = \frac{f_2^*(x_1, x_2, x_3)}{f_1^*(x_1, x_2, x_3)} \quad (30)$$

Here

$f_2^*(x_1, x_2, x_3)$ -relative values for the structural parameters of the secondary winding.

$f_1^*(x_1, x_2, x_3)$ -Relative Value of Efficiency Index

The relative value of the structural parameters of the secondary winding is calculated to dividing method the reference value by the actual value ($L_\sigma \cdot C_d$ maximum allowable value).

$$f_2^*(x_1, x_2, x_3) = \frac{f_2(x_1, x_2, x_3)}{C_{\max}}$$

Here

$C_{\max}$ -maximum permissible value of $L_\sigma \cdot C_d$

The efficiency is a relative value, so applies as it is.

(2) Value range of design variables

Considering the characteristics of the pulse transformer, the change range of design variables is chosen as follows.

- Current density $\Delta = 2.0 \sim 5.0$, A/mm².
- Flux density $B_{\text{iron}} = (0.3 \sim 1.2)$, T
- Shape coefficient. $\beta = (0.7 \sim 2.5)$.

(3) Constraint

- Temperature rising

The temperature rising is directly affected by the design variables.

Therefore, the main constraint is the temperature rising.

Considering the cooling condition of the pulse transformer and the allowable temperature rising of the winding by air is chosen to be 45 °C.

- no-load current

The no-load current is greatly affected by the flux density and the shape coefficient.

Considering the design plant voltage and capacity, the allowable value is chosen as 1.5%.

- The height of the iron core bridge

The height of the iron core bridge is chosen to be 120 mm, considering the manufacturing conditions of the ferrite core.

2) Solution of optimal design problem based on genetic algorithm

(1) Determination of the Fitness function

Considering that the optimal direction of the objective function is immediately correspond to the direction of the increase of fitness, the fitness function is written as follows.

$$F(t) = K_{\max} - \frac{f_2^*(x_1, x_2, x_3)}{f_1^*(x_1, x_2, x_3)}; \quad \max \quad (31)$$

In Equation (31), $K_{\max}$ is a sufficiently large parameter, which is determined by the degree of convergence of the genetic solution.

(2) Genetic operator design

The colony size was set in $n = 20 \sim 200$ range.

The selection operator makes a careful selection.

As a crossover operator, we take the form of a two-point crossover.

The mutation operator takes the standard mutation operator.

(3) Implementation of genetic algorithm by MATLAB

The above-designed genetic algorithm was simulated by MATLAB.

In the Start menu of the program, select Toolboxes → Optimization → Optimization Tool.

First, the fitness function is written in the TRANCFORMER.m file.

Function Trancformer (Id , B_f , beta)

---designed formula---

%-----fitness function-----

Fmax = Kmax-(Kcs/Cmax)/Eta1

if (io1<Iop) and (th_ago<th_k) and (L_f1<Lf1)

bool1=0;

In the program, Iop is the allowable value of no-load current, th_k is the allowable value of temperature rising, and Lf1 is the allowable value of core bridge height.

Then, the following operations are performed to simulate the designed genetic algorithm :

- Set the Solver option to ga-Genetic Algorithm in the Optimization Tool dialog box.
- Enter @my_fun in the Fitness function item.
- Enter 3 in Number of variables.
- Give the lower limit [20.30.7] and the upper limit [51.2.5] in the boundary condition (Bounds) item.
- Set to 100 the size of the population (Population size).
- Selection method is set as roulette method, mating is two-point mating.

Finally, set the “Use random states from previous run” item and click Start.

The variation character of the objective function F_x with the shape coefficient β and flux density B_f is shown in Figure 5 when current density is $\Delta=1.62$ A/mm², the variation character of efficiency η is shown in Figure 6, the variation character of $L_\sigma \cdot C_d$ size C is shown in Figure 7, and the variation character of the no-load current I_0 is shown in Figure 8.

When the shape coefficient β increases, the efficiency, cost and no-load current increase and the core leg height decreases.

If the flux density increases, the efficiency is not great change, the cost decreases and the no-load current increases.

If current density increases, the efficiency and $L_\sigma \cdot C_d$ decrease and the winding temperature rises.

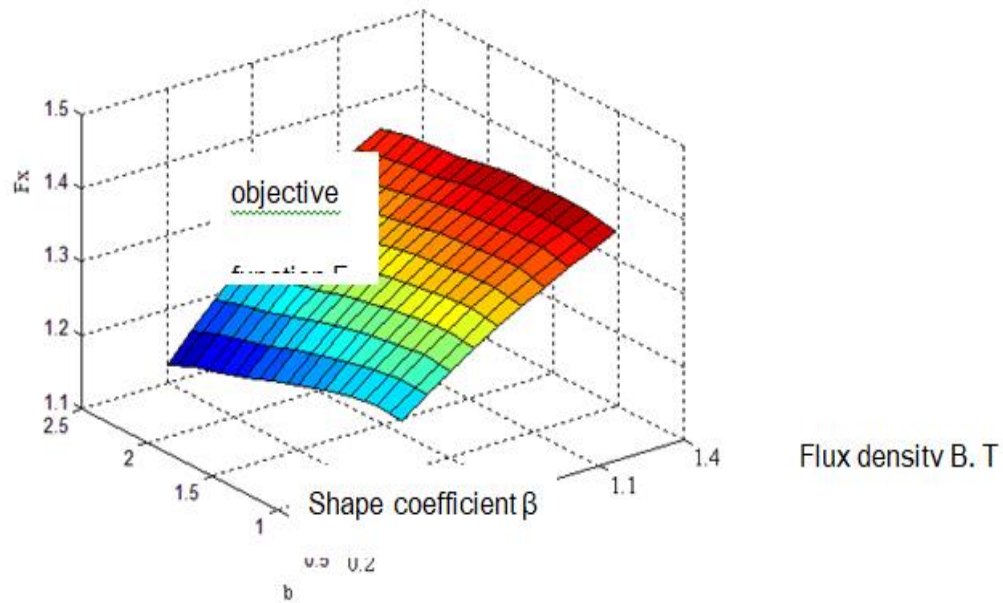


Figure 5. Variation character of objective function F_x by β and B_f

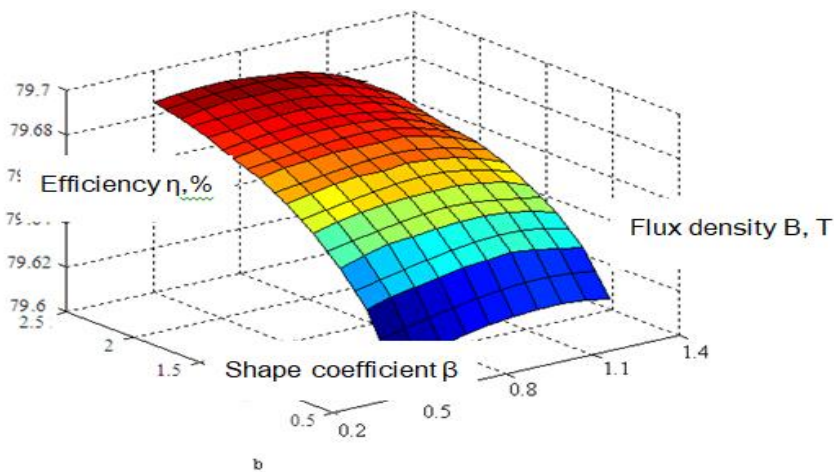


Figure 6. Variation character of efficiency η by β and B_f

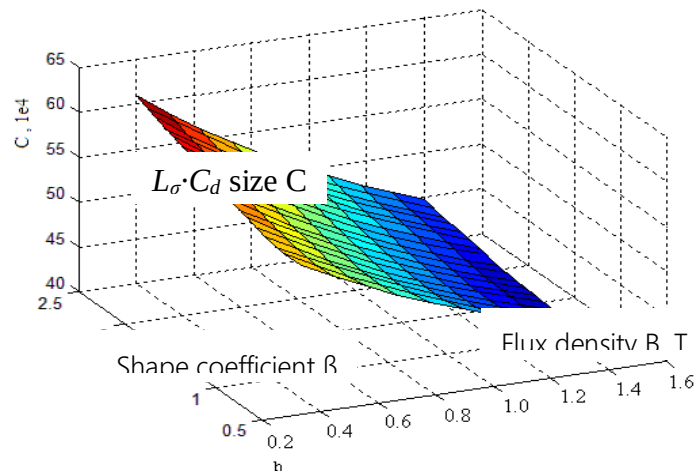


Figure 7. Variation character of the $L_{\sigma} \cdot C_d$ size C by β and B_f

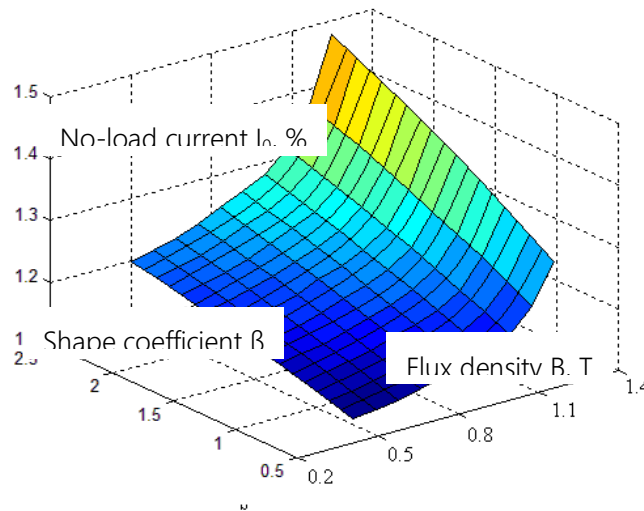


Figure 8. Variation character of no-load current I_0 by β and B_f

Optimal design and results analysis of pulse transformer

The technical character of the pulse transformer to be designed is as follows.

The output power is 15 kVA, operating frequency is 20 kHz, input voltage is 350 V, and output voltage is 15 kV.

The rise time of the pulse depends on the transformation ratio, so the output voltage of 15 kV is obtained by method that three pulse transformers are connected in series.

Therefore, the output voltage of the designed pulse transformer is 5 kV.

1) Optimal design of pulse transformer

The design variables obtained by applying the genetic algorithm are shown in Table 1.

Table 1. Optimal design variables

No	Design variable	Size
1	Flux density of the core (T)	0.32
2	Average current density of windings (A/mm ²)	1.62
3	Shape coefficient	2.2

When the operating frequency is 20kHz, the penetration depth due to skin effect is 0.467 mm, so the primary winding is manufactured by using copper wire that diameter is 0.9mm.

Table 2 shows the design results of the pulse transformer.

Table 2. Optimal design results of pulse transformers

No	Design parameter	Design value
1	Core cross-section (cm ²)	14.2
2	Window height (mm)	112
3	Window width (mm)	51
4	Number of primary turns (time)	15
5	Number secondary turn (time)	217
6	no-load current (%)	1.4
7	short-circuit voltage (%)	31.5
8	core mass (kg)	18.5
9	copper mass (kg)	2.4

10	Loss (W)	356
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As shown in Table 2, efficiency in one pulse transformer is 92.8% as the loss is 156 W.

2) Results analysis

Figure 9 shows a prototype of the designed and fabricated pulse transformer.

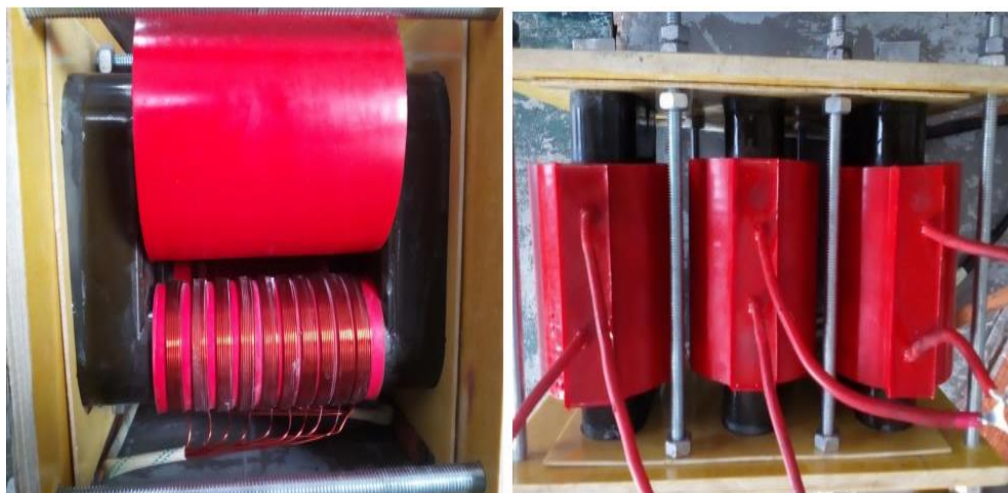


Figure 9. Prototype of Pulse transformer

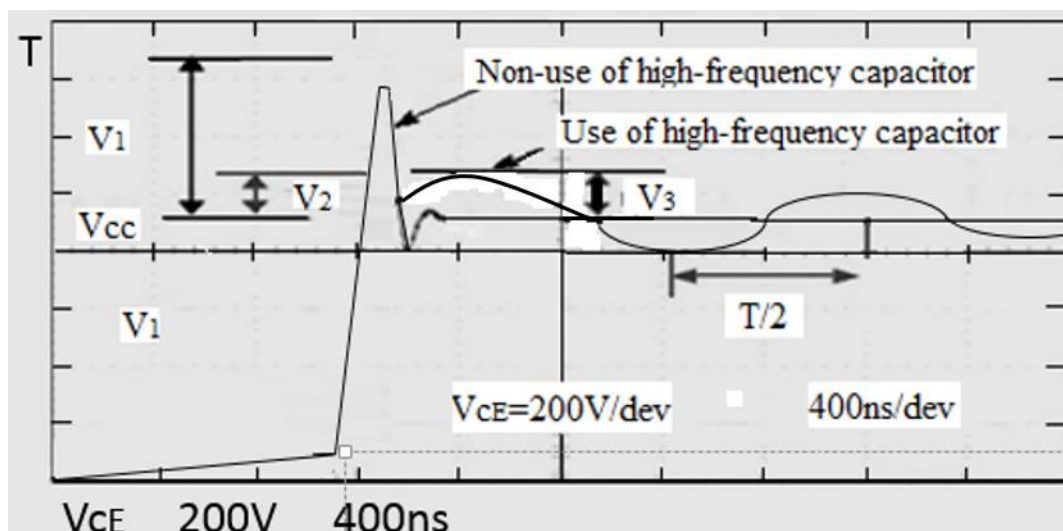


Figure 10 shows the waveform of the input pulse.

As shown in Figure 10, in case of high-frequency capacitor is not used, it leads to an increase of overshoot.

IV. Conclusions

In most cases of, the pulse transformer must transmit the shape of the pulse without distortion.

In the pulse transformers in use, the pulse duration is several microseconds.

In this context, the inductive reactance and effective resistance of the winding, the parasitic capacitance between the transformer core and the winding and between the individual parts of the winding can distort the shape of the pulses being transformed.

In this paper, an analysis on the dependence of the electromagnetic parameters according to structural dimensions of the pulse transformer to reduce the rise time of the pulse and a method to obtain the optimum result is suggested.

Based on the transformer model, it could be seen that the rise time of transformers is proportional to the product of the leakage inductance L_{σ} and the parasitic output capacitance C_d of the pulse transformer.

This product is calculated for three different transformer topologies: parallel, cone and foil winding concepts and it is shown that with a cone winding the fastest rise time can be achieved.

The resulting overshoot is defined by the $L_o C_d$ -ratio.

For the calculation of these parasitics an improved calculation procedure is proposed and validated by measurements.

And this paper is verified the correctness of the design results by analyzing the relationship between the influence of core material and switching speed and the Klystron (electron tube) with nonlinear inductance characteristics.

In the future, character analysis of pulse transformer by finite element method(FEM) and genetic algorithm will more correct and practical.

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