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RENESAS

DESIGN OF PUSH-PULL TYPE SWITCHING REGULATORS (BASIC)

1. INTRODUCTION

Regulator IC's such as the μPC305C have allowed simplification of circuitry and improvement in performance of series regulators. Little progress has been made in reducing the size and weight of these systems, however, due to large and heavy power transformers, smoothing capacitors and heat sinks.

In recent years highly efficient pulse width controlled power supplies (switching regulators), which are smaller and lighter, have been developed and are being used commercially, mainly for computers and terminal equipments. These switching regulators are gradually dominating applications in power supply systems.

Here we present a brief description of switching regulator designs from the standpoint of high voltage, high speed switching power transistors NTC1863-NTC1871A.

2. PRINCIPLE OF SWITCHING REGULATORS.

2-1 Explanation of Switching Regulators.

A choke input type smoothing circuit is shown in Fig. 1.

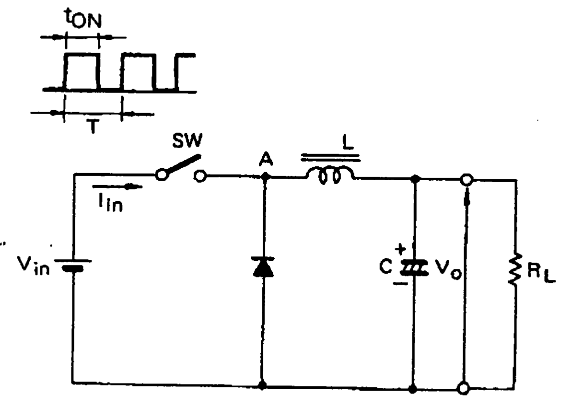

Fig. 1 (a) Choke Input Type Smoothing Circuit

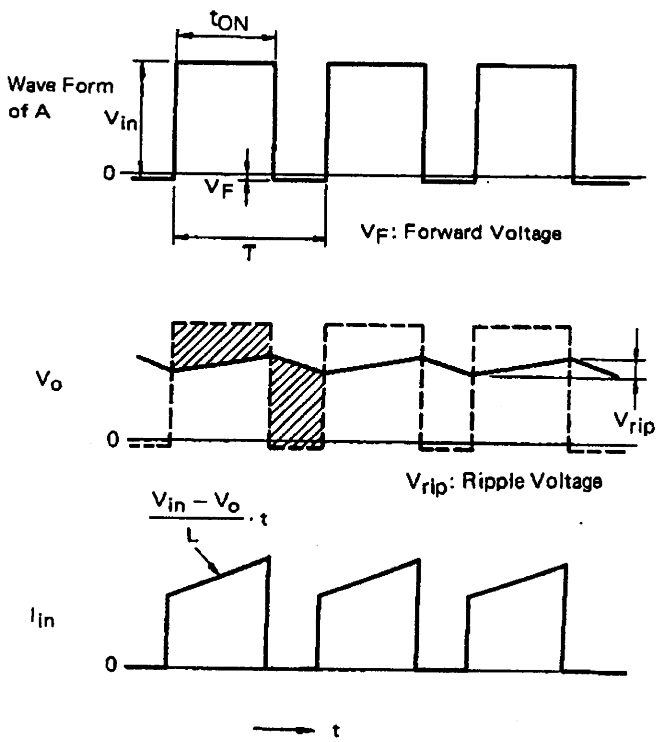

Fig. 1 (b) Operating Waveforms of Choke Input Type Smoothing Circuit

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In Fig. 1 (a), when switch SW is switched on-off with a period of T and an on-time of t_{ON}, the output voltage V_O will be approximately equal to the product of the duty cycle $D = t_{ON}/T$ of the on-time and the input voltage V_{in}.
That is:

$$V_O \fallingdotseq \frac{t_{ON}}{T} \cdot V_{in} \quad \cdots\cdots\cdots\cdots (1) \qquad \text{Provided that: } T \ll 2\pi\sqrt{LC}$$

$$R_L \leqq \frac{2L}{T - t_{ON}}$$

(conditions for current flowing through choke coil to be continuous)

From the above it can be seen that in order to obtain a constant output voltage V_O, regardless of the variations of the unregulated input voltage V_{in}, it is necessary that the duty cycle be changed according to variations of the input voltage.

Now, if the goal is to stabilize the output by reducing the pulse width t_{ON} to $t_{ON} - \Delta t_{ON}$, when the input voltage V_{in} increase to $V_{in} + \Delta V_{in}$, then,

$$-\Delta t_{ON} = -\frac{\Delta V_{in}}{V_{in} + \Delta V_{in}} \cdot t_{ON} \fallingdotseq -\frac{\Delta V_{in}}{V_{in}} \cdot t_{ON} \quad \cdots\cdots\cdots\cdots (2) \qquad \text{Provided that: } \Delta V_{in} \ll V_{in}$$

Therefore, it can be seen that it is sufficient if the pulse width is varied an amount approximately proportional to the variation of the input voltage ΔV_{in}.

2-2 Operating Principle of Switching Regulators.

A simple switching regulator circuit is shown in Fig. 2.

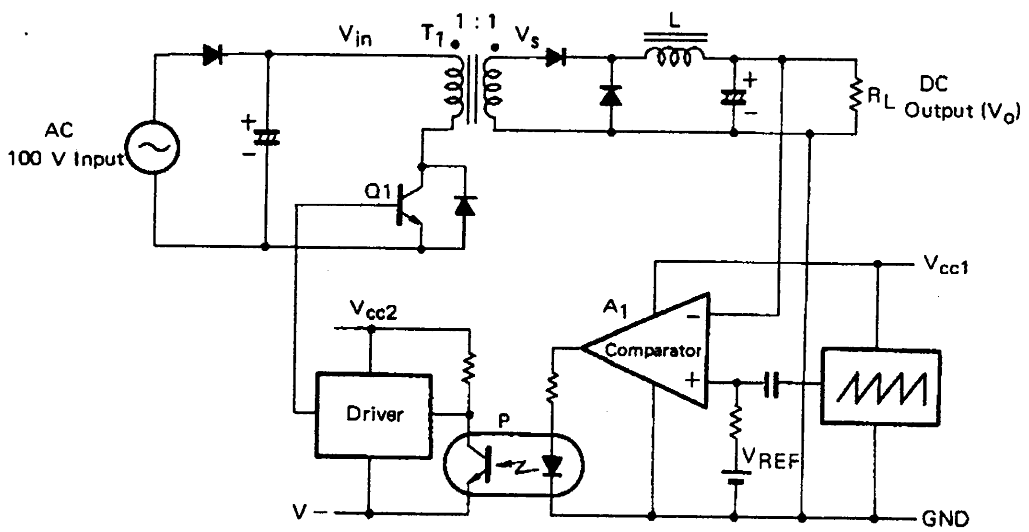


Fig. 2 Single-Ended Operation Type Regulator Circuit

Generally, since power supply units are generally used in a floating operation, the input and output are isolated by a high frequency transformer T_1.

Due to the switching of transistor Q_1, the primary winding of T_1 is excited and, since the secondary winding has the same number of turns as the primary winding, a voltage $V_s = V_{in}$ is generated in the secondary winding. The output circuit is the same type of choke input smoothing circuit as shown in Fig. 1, therefore, the output voltage V_o is given by the equation (1). In order to stabilize the output voltage, the variation of output voltage is sensed and fed back to control the driving pulse width. A_1 is a comparator having the output voltage consisting of a sawtooth wave superposed upon the reference voltage applied to the non-inverting input terminal V_N. Therefore, since the slice level of the sawtooth wave at the output of the comparator will change according to the slightest variation of the output voltage V_o, the output of A_1 will become a pulse width modulated signal having the same period as that of the sawtooth wave. When V_o increases this signal acts to decrease the width of the pulse, and enters the driver through photo-coupler P to drive transistor Q_1. As described above, the on-time of the transistor is controlled according to the variations of the output voltage, and a stable output is always obtained.

3. KINDS OF SWITCHING REGULATOR CIRCUITS AND OPTIMUM SWITCHING POWER TRANSISTORS.

NEC's optimum switching power transistors for switching regulators are shown in Table 1.

Table 1. Product Series of High Speed Switching Power Transistor

Outline / $I_{C(DC)}$ / P_T* / V_{CEO}	Type No.		
	TO-66 7 A 50 W	TO-3 7 A 100 W	TO-3 10 A 150 W
100 V	NTC1863	NTC1866	NTC1869
250 V	NTC1864	NTC1867	NTC1870
400 V	NTC1865	NTC1868	NTC1871A

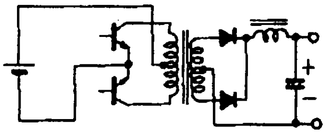
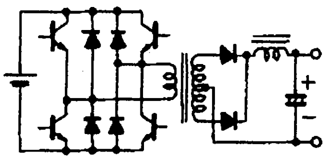
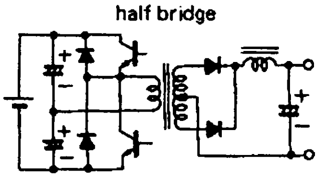
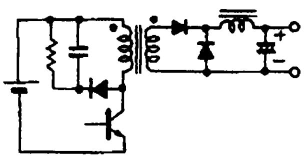
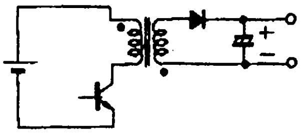
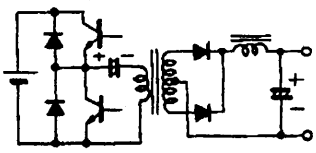
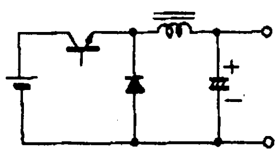
* Tc = 25 °C

The above table shows a series of high voltage, high speed switching power transistors newly developed for use in switching regulators, in which, compared to usual power transistors, an extremely high switching speed of t_{ON}, $t_f < 1$ μs and $t_{stg} < 2$ μs has been realized.

Various circuits of switching regulators, and the optimum switching power transistors are shown in Table 2.

In addition to the single-ended operation type described previously, there are various switching regulator circuits of such as the push-pull type, bridge type, etc. And, since the voltage range and current capacity required to the switching element will vary according to the input voltage and output capacity of the regulator, it is necessary to select the optimum transistor from Table 1. Among the various circuits, the push-pull type switching regulator as compared with the single-ended operation type has a high utilization factor of the high frequency transformer and secondary side smoothing circuit; and compared with the bridge type requires only half the number of switching elements, thus simplifying the driver circuit. For these reasons, the push-pull type is most frequently used as a switching regulator.

Table 2. Circuits of Switching Regulators and Optimum Switching Power Transistors

Circuits		Optimum Power Transistor — Input Voltage: AC 100 V or DC 130 V	AC 220 V	DC 12 V or 24 V	DC 48 V	Output Capacity (W)
Types of Inverter	push-pull 	NTC1865	V_{CEO}=900 V class Po. Tr.	NTC1863 or NTC1866	NTC1864	100
		NTC1868	"	NTC1869	NTC1867	200
		NTC1871A	"	–	NTC1870	~500
	full bridge 	NTC1864	NTC1865	NTC1863 or NTC1866	NTC1863	150
		NTC1867	NTC1868	NTC1869	NTC1866	250
		NTC1870	NTC1871A	–	NTC1869	500
	half bridge 	NTC1864	NTC1865	–	NTC1863 or NTC1866	75
		NTC1867	NTC1868	–	NTC1869	150
		NTC1870	NTC1871A	–	–	300
	single-ended (ON-ON) 	NTC1865	V_{CEO}=900 V class Po. Tr.	NTC1863 or NTC1866	NTC1864	50
		NTC1868	"	NTC1869	NTC1867 or NTC1870	100
		NTC1871A	"	–	–	200
	single-ended (ON-OFF) 	NTC1865	"	NTC1863 or NTC1866	NTC1864	50
		NTC1868	"	NTC1869	NTC1867 or NTC1870	100
		NTC1871A	"	–	–	200
	single-ended push-pull 	NTC1864	NTC1865	NTC1863 or NTC1866	NTC1863 or NTC1866	75
		NTC1867	NTC1868	NTC1869	NTC1869	150
		NTC1870	NTC1871A	–	–	300
chopper type 		(NTC1864)	(NTC1865)	NTC1863	NTC1863	50
		(NTC1867)	(NTC1868)	NTC1866	NTC1866	75
		(NTC1870)	(NTC1871A)	NTC1869	NTC1869	150

4. DESIGN OF PUSH-PULL TYPE SWITCHING REGULATORS.

We will now describe a typical method of designing push-pull type switching regulators.

4-1 Composition.

The block diagram of push-pull type switching regulator is shown in Fig. 3, and the timing chart of the circuit is shown if Fig. 4. The circuit of Fig. 3 differs from the single-ended switching regulator in the point that a separately excited high frequency push-pull inverter is used for voltage conversion. Since in the circuit of Fig. 2 the output voltage is directly compared with the sawtooth wave, the loop gain should be low and allowing large variations of the output, so an error amplifier of about 50 dB gain is necessary as shown in Fig. 3. In addition, in order to provide the output signal into 2 phases is required in an after the isolator; and also an over-current flowing in the switching elements during overloads. Also, auxiliary power source is required to supply power to these control circuits. This power source consists of a line operated DC–DC converter, in order to permit AC/DC operation and reduction of size.

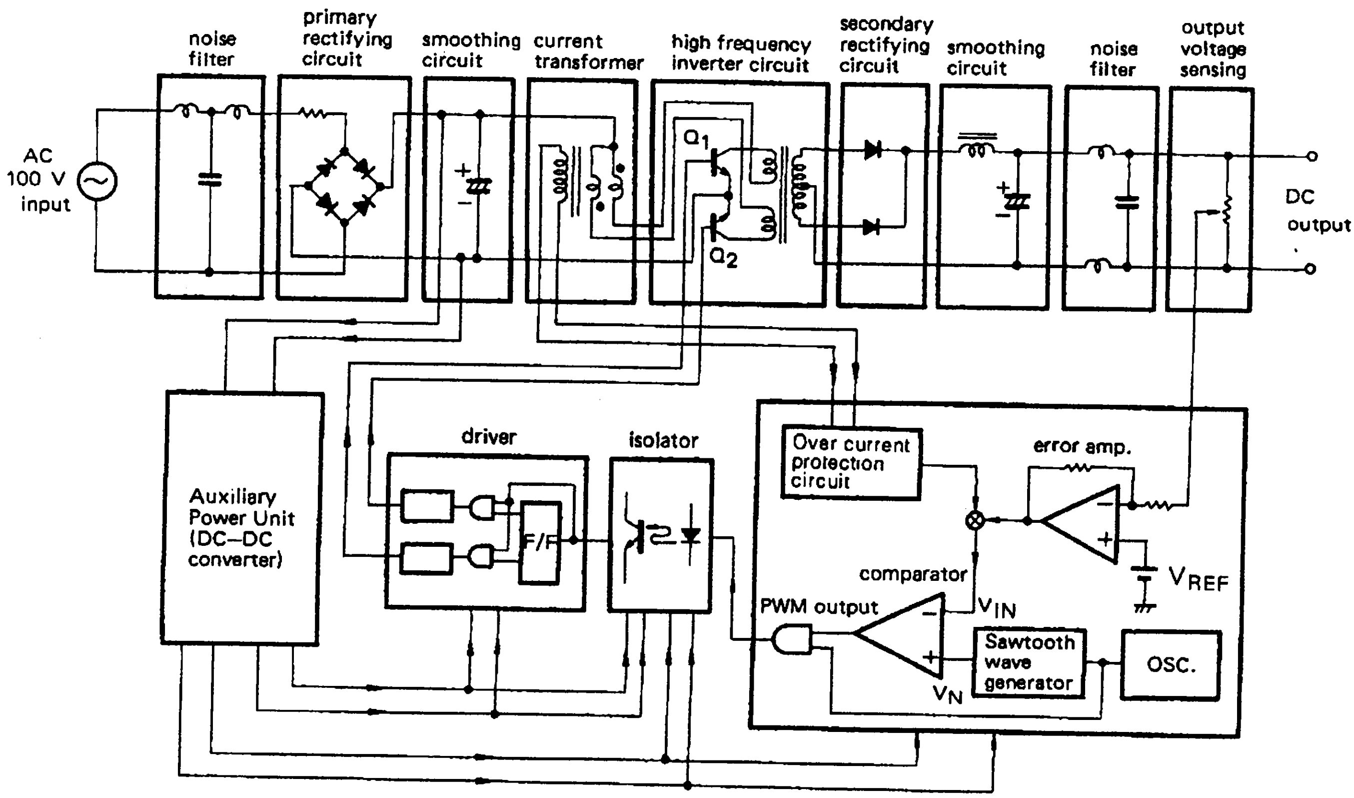


Fig. 3 Block Diagram of Push-Pull Type Switching Regulator.

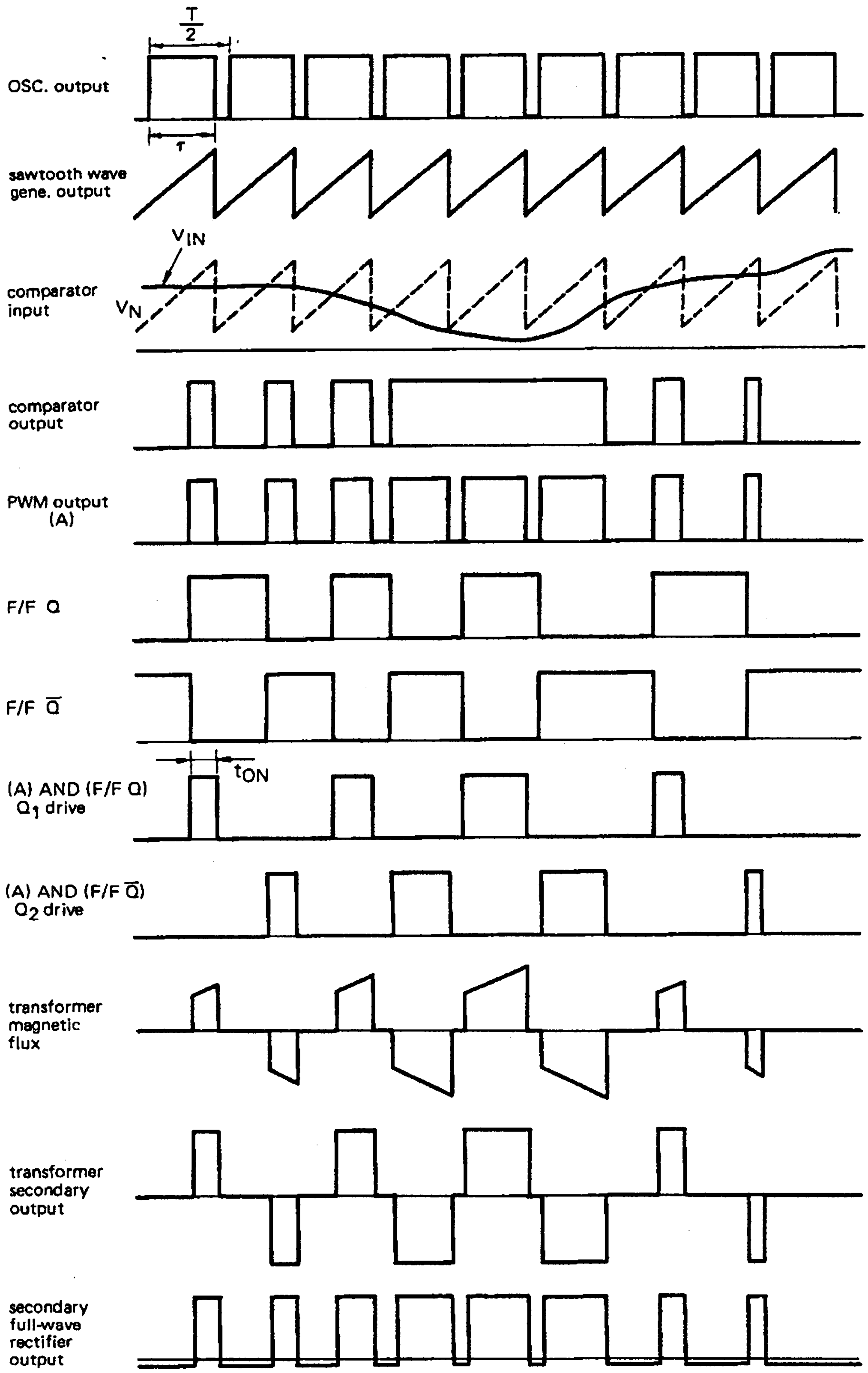


Fig. 4 Timing Chart of Push-Pull Type Switching Regulator Circuit

4-2 Design Method of the Various Blocks.

(1) Primary Rectifying and Smoothing Circuit.

A line operated full-wave rectifying and smoothing circuit is composed of a bridge rectifier and a smoothing capacitor, and the rush current flowing into the rectifier and capacitor when power is turned on is limited by a low resistor in series with the AC line. Also, in the case of large capacity power supplied, slow starting circuits using SCRs or relays may be employed.

The capacity of the smoothing circuit is determined by the input capacity of the power unit and the required ripple level, and for this purpose there is an extremely convenient curve (chart) by O. H. Schade, as shown in Fig. 5.

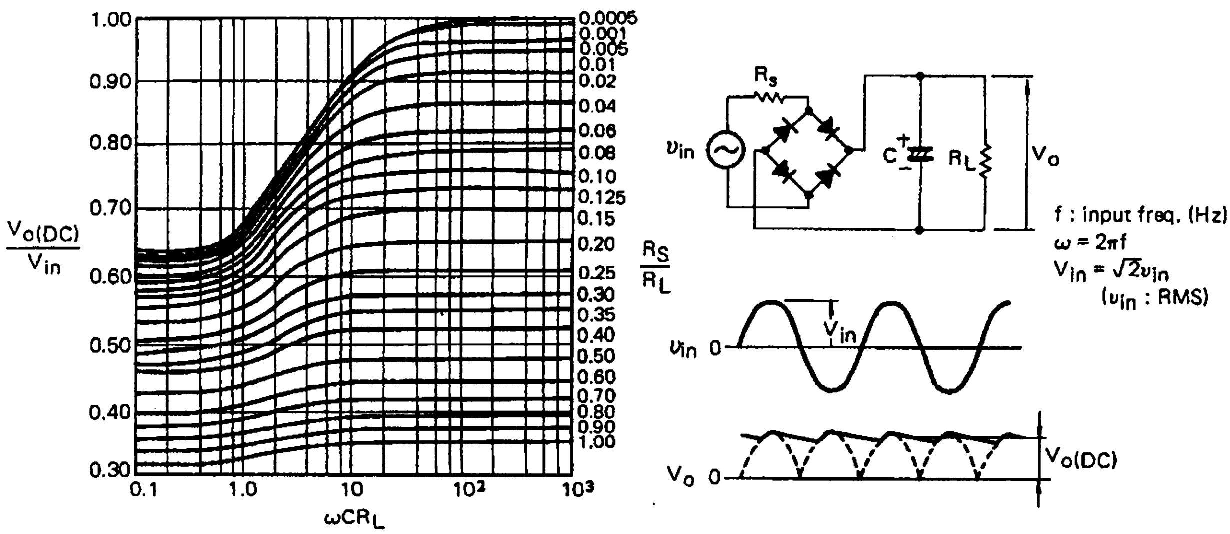


Fig. 5 O. H. Schade's Curve (for determining Circuits Constants of Full-Wave Rectifying Circuits.)

(2) Design of Transformers.

In order to reduce size and avoide audible noise, high frequency is used in switching regulators. However, considering the high frequency characteristics of diodes, transistors, capacitors, etc. a frequency of about 20 kHz is generally used in push-pull type regulators.

i) Material and Shape of Cores.

From the point of saturation (magnetic) flux density, permeability, and iron loss Tohoku Kinzoku's 3100B or TDK's H_{5A} is used as the core materials of transformers and choke coils.

EI shaped cores are easy to handle. The volume of the core is determined from such factors as follows: the input voltage, the number of turns of the primary winding which is determined from the frequency used, the cross section area of the primary winding which is determined from the output capacity, the primary winding inductance, and the superposed DC current.

When a square wave voltage of frequency f (Hz) and peak value V_{in} (V) is applied to a transformer which has N_p turns as the primary winding around a core of A_e (m^2) cross section, the relationship between the flux density B (wb/m^2) created in the core and V_{in} can be shown by the following expression.

$$V_{in} = 4 \cdot B \cdot A_e \cdot f \cdot N_p \text{ (V)} \quad \cdots\cdots (3)$$

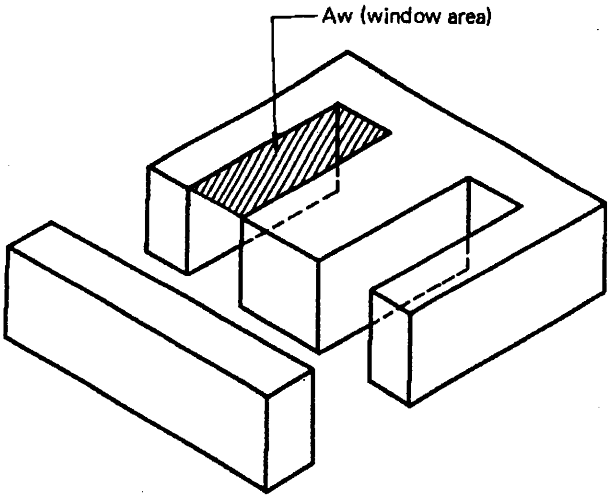


Fig. 6 Shape of Core

On the other hand, if in Fig. 6 δ (A/m^2) is the allowable current density of the winding, and w.f. (window factor) the proportion (space factor) of the total cross sectional area of the primary and secondary windings (A/m^2) to the total window area $A_W(m^2)$ of the core, then,

$$A_C = (w.f.) \cdot Aw = \frac{2N_P \cdot I_P}{\delta} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (4)$$

Therefore, from equations (3) and (4) the maximum input P(VA) which can be applied to the transformer is:

$$P = V_{in}.I_P = 2B.f.A_e.(w.f.).\ A_W.\ \delta\ (VA) \quad \ldots\ldots\ldots\ldots\ldots (5)$$

Now, when in equation (5) the following values are substitued for the items below, equation (6) is obtained.

$B = 1500$ (Gauss) $= 1.5 \times 10^{-1}$ (wb/m^2)

(Generally, taking in consideration the balance of the decrease of saturation (magnetic) flux density with rise in temperature and iron loss, a magnetic flux density of 1500 Gauss is used.)

$\delta = 2(A/mm^2) = 2 \times 10^6$ (A/m^2)

(From the temperature rise caused by the specific resistance of the copper wire, generally, 2 A/mm^2 is used as the allowable current density of the copper wire employed in transformer windings.)

w.f. = 0.7

(From the thickness of the insulating material of the windings and the space taken up by the shield, 0.7 is generally used for the w.f.)

f = 20 (kHz)

then,

$$P = 8.4 \times A_e \times A_W \times 10^9\ (VA) \quad \ldots\ldots\ldots\ldots\ldots\ldots (6)$$

When the maximum input capacity P(VA) of Tohoku Kinzoku's EI shaped cores are determined using equation (6) the results are as shown in Table 3. Actually, when employing these cores in switching regulators, due to the necessity of taking into consideration a margin of the (magnetic) flux density to cope with variations of the input and also the DC superposed characteristics, it is common for the design to have a considerable allowance in regards to the P(VA) values of Table 3.

Table 3. Kinds & Maximum Input Capacity of EI Cores.

$B = 1500$ Gauss, $f = 20$ kHz, $\delta = 2$ A/mm², w.f. = 0.7 * Core material: 3100B

Kind of core	Total window area of core A_W (cm²)	Effective cross section area of core A_e (cm²)	Maximum input capacity P (VA)	Practical example of output capacity P_o (W)
FEI 30	1.44	1.09	132	25
FEI 40	3.10	1.46	380	50
FEI 50	4.78	2.30	923	100 – 150
FEI 60	7.84	2.45	1,600	Undetermined

ii) Windings.

(a) Number of turns of primary winding: N_p

N_p is determined, employing equation (3), so that the design (magnetic) flux density B shall be obtained at maximum input voltage $V_{in(max)}$.

$$N_p = \frac{V_{in(max.)}}{4 \cdot B \cdot A_e \cdot f} \qquad (7)$$

(b) Diameter D_p of primary winding wire.

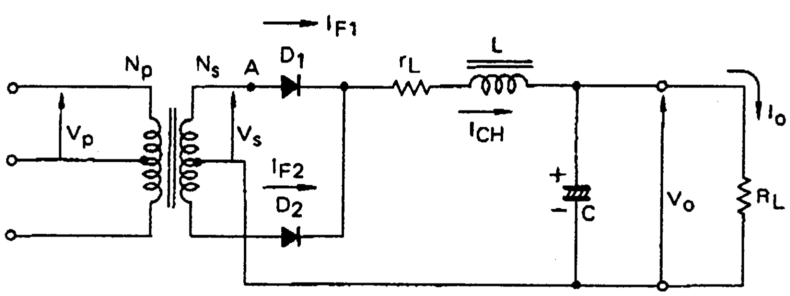
If an average current through into the primary winding is $I_{c(av)}$.

$$\frac{I_{c(av)}}{\pi \cdot \left(\frac{D_p}{2}\right)^2} \leqq \delta$$

$$D_p \geqq \sqrt{\frac{4 I_{c(av)}}{\pi \delta}} \quad (m) \qquad (8)$$

(c) Number of turns of secondary winding: N_s

N_s is designed so that the specified output voltage will be obtained under the worst conditions (minimum input voltage, maximum load, and minimum ambient temperature).



Vp: primary input voltage (V)
Vs: secondary output voltage (V)
Vo: smoothed output voltage (V)
VF: forward voltage (V) of diodes D1 and D2
L: inductance (H) of choke coil
rL: internal resistance (Ω) of choke coil
C: capacity (F) of smoothing capacitor
RL: load resistance (Ω)
Io: output current

Fig. 7 Secondary Smoothing Circuit

In the secondary smoothing circuit shown in Fig. 7, the relationship between the input and output voltages and number of turns of the primary and secondary windings of the transformer is shown by the following equation.

$$V_S = \frac{N_S}{N_P} \cdot V_P \quad (V) \quad \cdots\cdots (9)$$

And, when an output voltage of period T(sec), maximum pulse width τ (sec), and peak value V_S (V) is induced in the secondary side of the transformer, the output voltage V_O(V) obtained when a load current I_O(A) is taken out is shown by the following equation.

$$V_O = V_S \cdot \frac{\tau}{\frac{T}{2}} - V_F - I_O \cdot r_L \quad (V) \quad \cdots\cdots (10)$$

Therefore, from equations (9) and (10) the number of turns of the secondary winding can be determined from the following equation.

$$N_S \geqq \frac{(V_O + V_F + I_O \cdot r_L) \cdot N_P \cdot T}{2\,\tau \cdot V_P} \quad \text{(turns)} \quad \cdots\cdots (11)$$

(d) Cross section of secondary winding.

When round copper wire is used for the secondary winding, the same as for the primary winding, the diameter D_S(m) of the secondary winding can be determined from the following equation, similarly to equation (8).

$$D_S = \sqrt{\frac{4\, I_O(\text{max.})}{\pi\, \delta}} \quad (m) \quad \cdots\cdots (12)$$

Provided that, I_O(max): Maximum output current (A)

However, in the case of low output voltage, large current capacity power such as are used in terminal equipment, foil may be used for the winding to reduce skin effect. In this case the cross sectional area S_S(m^2) of the winding is determined from the following equation.

$$S_S = \frac{I_O(\text{max.})}{\delta} \quad (m^2) \quad \cdots\cdots (13)$$

iii) Static shield.

To prevent the influence of noise to be easily passed through the static coupling due to the stray capacity between the primary and secondary windings of the transformer, it is necessary to provide static shield by inserting metallic foil between the primary and secondary windings, or by other means.

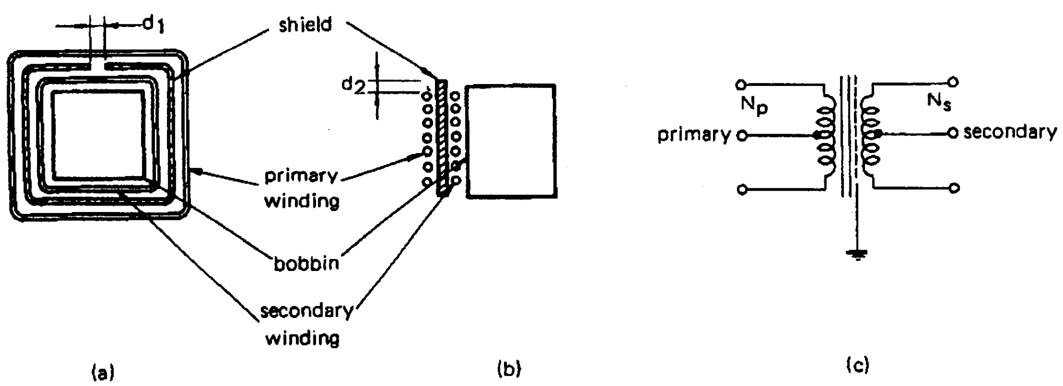


Fig. 8 Static Shield of Transformer

Shielding of the transformer is conducted as shown in Fig. 8, but care should be taken since when d_1 of the shielding foil joint in Fig. (a) is too large the shielding effect will decrease and, when the joint overlaps, loss will increase. Also, unless the shield foil extends beyond the edges of the winding the static coupling at the edges of the windings cannot be decreased, therefore it is necessary to provide distance d_2 as shown in Fig. (b).

(3) Secondary Rectifying and Smoothing Circuit.

Fig. 9 shows the waveform during operation of the various point of the secondary smoothing circuit shown in Fig. 7 previously.

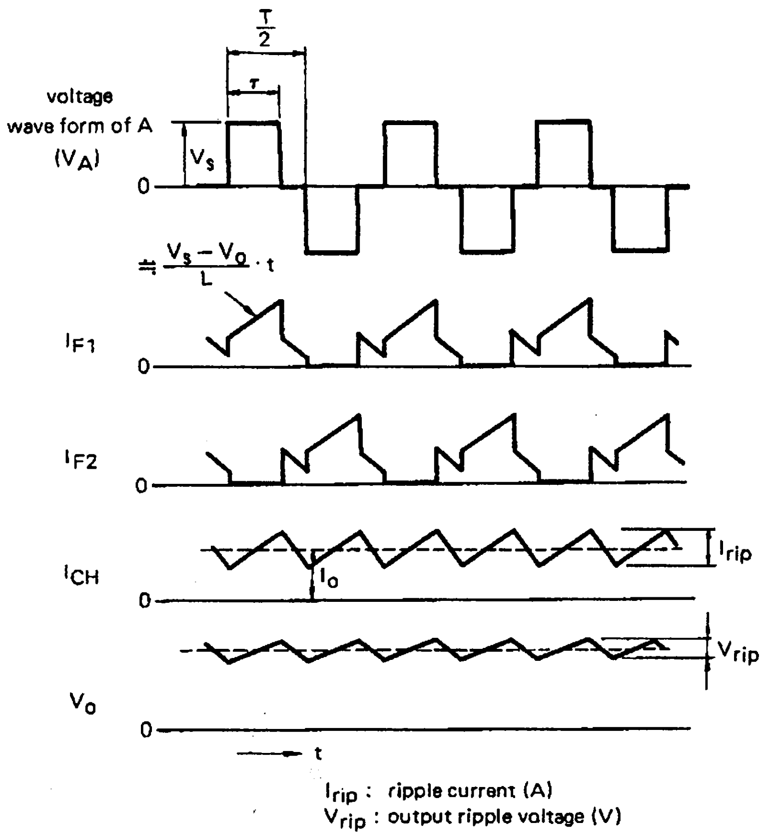


Fig. 9 Waveforms of the various parts of Secondary Smoothing Circuit

During interval τ (sec) of period T (sec) a voltage V_s (V) will be induced in the secondary of the transformer, and during this interval the smoothing circuit will be supplied with power from the primary. During this interval a current having a gradient of $(V_s - V_o) \cdot t/L$ will flow in the rectifier elements D_1 and D_2 connected to the terminals of the secondary, and this current will flow through the choke coil in a superposed manner. On the other hand, during the quiescent time $T/2 - \tau$ (sec) the choke coil will function to continue the current, thus rectifier elements D_1 and D_2 will act as flywheel diodes. As a result, the current flowing in the choke coil taked the form, as shown in Fig. 9, of a ripple current I_{rip} (A) superposed on a current equal to the output current I_o (A). This ripple current flows into the smoothing capacitor connected in parallel with the output terminals.

i) Selection of Rectifying Elements.

Since the currents I_{F1} and I_{F2} flowing in the rectifying elements D_1 and D_2 take the form shown in Fig. 9, it is necessary to select elements for D_1 and D_2 having a current capacity appropriate for I_{F1} and I_{F2}. Also, since in normal state a maximum voltage of $2V_s$ (V) is applied to these elements, it is necessary to select elements capable of withstanding this voltage. In addition, as the secondary rectifying circuit operates at a frequency of about 20 kHz, the use of fast recovery diodes having a short recovery time is required. Specifically, in the case of power units of low voltage (≃5 V), large current output the use of Schottky barrier diodes, which have an extremely fast recovery time, is appropriate.

ii) Smoothing Capacitors.

As described earlier, the secondary smoothing circuit is of the choke input type, and the smoothing factor ($\alpha = V_{rip}/V_s$) of such a circuit is generally shown by the following equation.

$$\alpha = \frac{V_{rip.}}{V_s} = \frac{1}{\omega^2 LC} \quad \ldots\ldots (14)$$

$$\omega = 2\pi f$$

However, since electrolytic capacitors, which do not have very good frequency characteristics, are employed for smoothing, the impedance Z_C of the capacitors cannot be expressed as $Z_C = 1/\omega C$ at a frequency around 20 kHz. The smoothing factor is usually determined from the following expression.

$$\alpha = \frac{Z_C}{\omega L} \quad \ldots\ldots (15)$$

Z_C: Impedance of capacitor at frequency used.

iii) Choke Coils.

As described before, a ripple current I_{rip} inversely proportional to the inductance L of the choke coil will flow in the choke coil. But, when the load is light and the output current I_o becomes less than $2\,I_{rip}$, the current flowing in the choke coil will become discontinuous and the filtering effect will be lost. This value is called the critical current I_{CHO} (A), and expressed as follows.

$$I_{CHO} = \frac{V_s - V_O}{2L} \cdot t_{ON} \ (A) \quad \ldots\ldots (16)$$

The critical current is related to the size of the dummy resistor used to prevent an abnormal rise of the output voltage, and is usually taken at 1/7 to 1/10 of the rated output current $I_{o(max)}$ (A).

that is,

$$I_{CHO} \leqq \left(\frac{1}{7} \sim \frac{1}{10}\right) \cdot I_{omax.} \ (A) \quad \ldots\ldots (17)$$

Therefore, from equations (1), (16) and (17) the inductance required by the choke coil will become:

$$L \geqq \frac{(3.5 \sim 5) \cdot (V_s - V_O)}{I_{omax.}} \cdot \frac{V_O}{V_s} \cdot T \ (H) \quad \ldots\ldots (18)$$

EI shaped Ferrite cores, similar to the for choke coils. But since direct current is superposed upon the choke coils, gapped cores are necessary to prevent (magnetic) flux saturation.

The number of turns and core volume required for a choke coil having the inductance and rated output current $I_{o(max)}$ (A) shown in equations.

The number of turns N_1 are shown by the well known in equation (20).

$$L = AL \times N^2 \times 10^{-9} \ (H) \quad \ldots\ldots (19)$$

AL: AL – value (nH/T^2) of core at gap t.

$$N_1 = \sqrt{\frac{L}{AL} \times 10^9} \quad \text{(turns)} \quad \ldots\ldots \quad (20)$$

Core volume is determined by selecting a core in which the product of the DC superposed current under actual load and the number of turns does not exceed the DC superposed characteristic NI (ampere-turns) of the core at the gap t (mm).

That is,

$$NI > N_1 \times I_{omax}. \quad \text{(Ampere Turns)} \quad \ldots\ldots \quad (21)$$

(4) Auxiliary Power Supply Circuits.

As shown in Fig. 10, all of the DC–DC converters used as auxiliary power supplies operate on the principle of having a base feedback winding on the transformer to sustain self-excited oscillation.

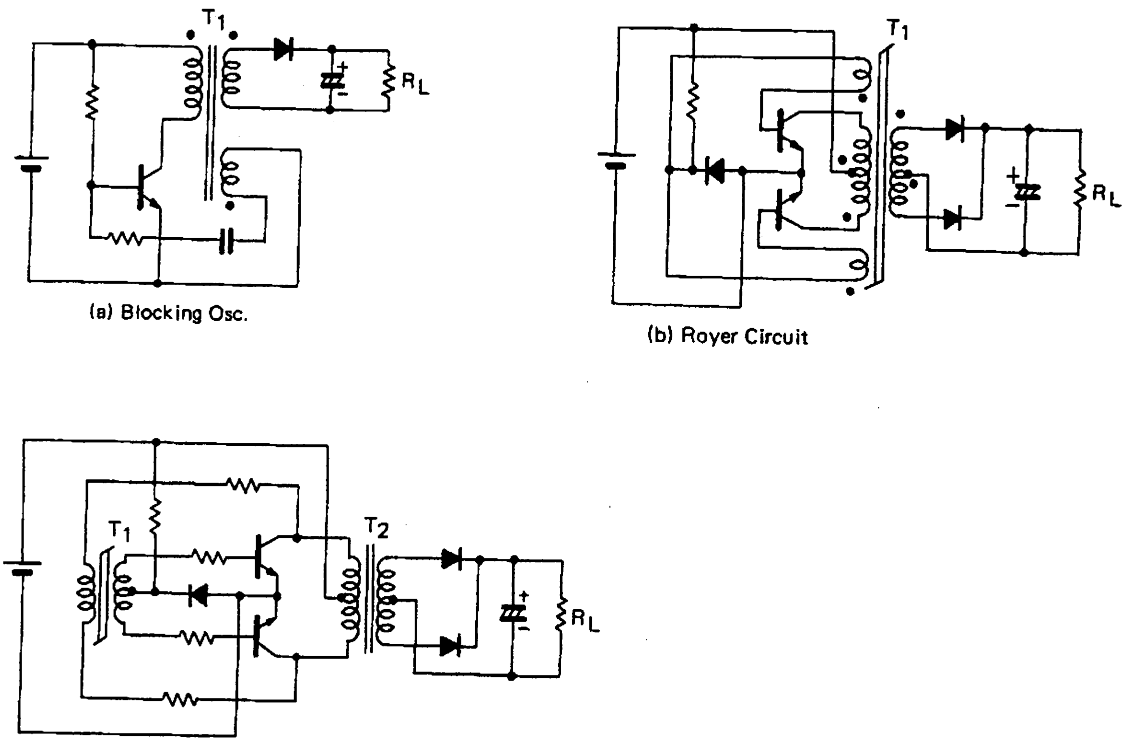


Fig. 10 Examples of DC–DC Converter Circuits for Auxiliary Power Supplies.

Among the circuits shown, when the output is small the Blocking Oscillator type shown in Fig. 10 (a) is suitable from the standpoint of needing a few parts. Design of transformers for converters is conducted in the same way as described in Item (2). But particularly, since the transformers T_1 of circuits (b) and (c) of Fig. 10 are used at saturated magnetic flux, TDK's H_{5C2}, which has a square hysteresis characteristic, is suitable as core material. Since converters are operated from the AC100 V – 117 V line, transistors of the V_{CEO} = 400 V are required for use in converters. When it is necessary to stabilize the output of converters, as in control circuits which are operated from the output of the auxiliary power supply, monolithic voltage regulators of the μPC143xxH or μPC78L are used.

(5) Base Driving Circuits.

The base driving circuit of the main switching transistor is important from the standpoint of reducing the switching time and thus the switching loss of the transistor. Particularly, since the high voltage, high speed switching transistor which is the main switching element has a low h_{FE} of about 10, the base current becomes considerably larger. Therefore, a pulse transformer is generally used to provide current amplification. (Sometime it also performs the function of isolation.)

A pulse transformer of 2 : 1 to 5 : 1 turns ratio is employed in the circuit of Fig. 11, and extraction of the carrier accumulated in the base of the final stage transistor is performed through diode D_1 by the backswing voltage generated in the pulse transformer during OFF time. (Since in this method the backswing voltage becomes low when the width of the input becomes narrow, some ingenuity is required in arranging the circuit.)

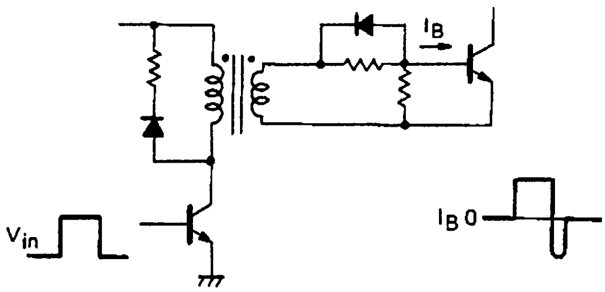
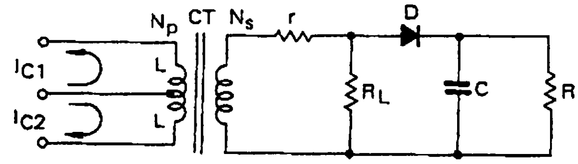


Fig. 11 Base Driving Circuit

(6) Overcurrent Sensing Circuit.

Current sensing methods for protecting the main switching transistor and primary and secondary rectifying diodes from overcurrent are: the method of utilizing the voltage drop of a low resistance resistor placed in series in the GND line of secondary circuit; and the method of directly sensing the collector current of the transistor employing a current transformer.

However, from the points of ease of control and low power loss in the sensing section the latter method is superior. A circuit using this method is shown in Fig. 12. The peak value of the collector current is detected by the diode D and capacitor C in the secondary of the current transformer (CT).



L : Inductance of primary winding wire.
CT: TDK's H_5C_2 T shaped core used.
$N_p = 3T$, $N_s = 150T$

Fig. 12 Overcurrent Sensing Circuit

i) Core Material and Shape.

In order to improve detecting sensitivity and to reduce sagging of the output waveform as much as possible, TDK's H_5C_2 toroidal cores which have high permeability and good coupling coefficients are used.

ii) Windings.

The number of turns of the primary and secondary also greatly effect the detection sensitivity, and a turn ratio of about 1 : 50 is usually employed. The sag D, shown in the output wave form of Fig. 13, is determined by the internal resistance r of the winding and the terminal resistor R_L. The sag D is expressed by the following equation.

$$D = \frac{r \cdot R_L}{R_L + r} \cdot \frac{\tau}{L} \qquad (22)$$

τ : Width of input pulse
L: Primary inductance of CT

Therefore, in order to reduce the sag, it is necessary to use a wire having larger diameter for the secondary winding to reduce the internal resistance.

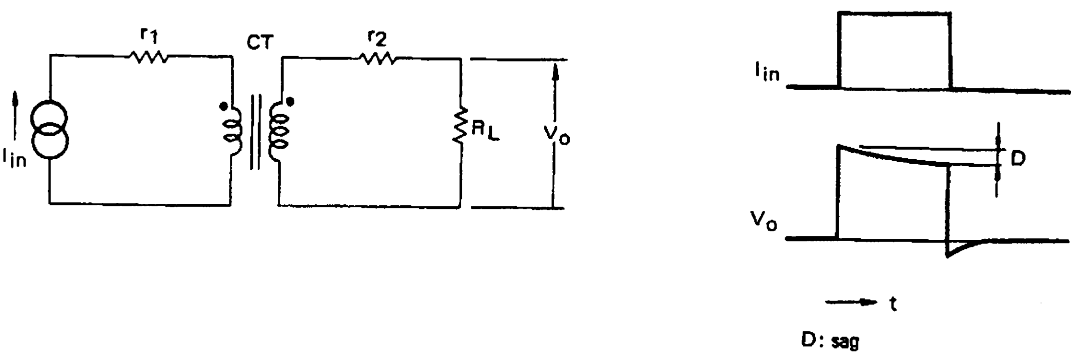


Fig. 13 Operating Waveforms of Current Transformer

5. ENDING

Design methods for switching regulators centering around the selection of the elements and design of the transformers have been briefly described above. Design of control circuits and actual examples of the design of switching regulators are described in the "Applications" edition, to which we refer you.

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