2.2 Design with Reference to the Modulation Procedures

2.2.1 Modulation Procedures

The switched-mode DC voltage regulator is a closed-action-chain system. The regulation process becomes an intermittent one by the conversion of the error signal into a sequence of pulses which act on the switching element of the energy transfer circuit.

The output signal of the comparator comparing the reference voltage with the pilot signal obtained from the output sampler divider, is amplified and passed to the input of the A–D converter. Depending on the value of amplified error signal applied to the input, some parameters of the pulse train at the output is varied (modulated). This pulse train drives the switching transistor (or transistors) of the energy-transfer circuit.

Two main categories of modulation are distinguished, depending on the particular parameter of the pulse train which is varied:

(1) *Pulse-time Modulation (PTM)*

In this modulation procedure the error signal modulates the time dependence of some parameter of the pulse train. This is also termed "Time Ratio Control" (TRC).

The amplitude of the pulse train is constant. Within this category, one of two different modulation procedures may be applied in the switched-mode regulator.

— *Pulse-width modulation*: a pulse-time modulation in which the error signal modulates the width of pulses. The pulse-width modulation is often referred to as "pulse-duration modulation". Their respective abbreviations are PWM (Pulse-width Modulation) and PDM (Pulse-duration Modulation). In the case of a pulse-width modulated signal, the temporal position of the leading or trailing edge of the pulse (or both) will vary with the instantaneous value of the modulating error signal. The pulse repetition rate and its reciprocal value, the cycle time $T = 1/f$, and the pulse amplitude are constant. Therefore, it is also referred to as "constant frequency TRC".

— *Pulse-frequency modulation* (PFM): a pulse-time modulation in which the error signal modulates the instantaneous frequency (i.e. the number per unit time) of the pulses. Therefore, it is also referred to as "variable frequency TRC".

(2) *Pulse Amplitude Modulation (PAM)*

In this procedure the amplitude of the pulse train is modulated. At the same time, other parameters of the pulse train (cycle time, duration, position) remain unchanged. Pulse-amplitude modulation is rarely applied in the technology of power supplies.

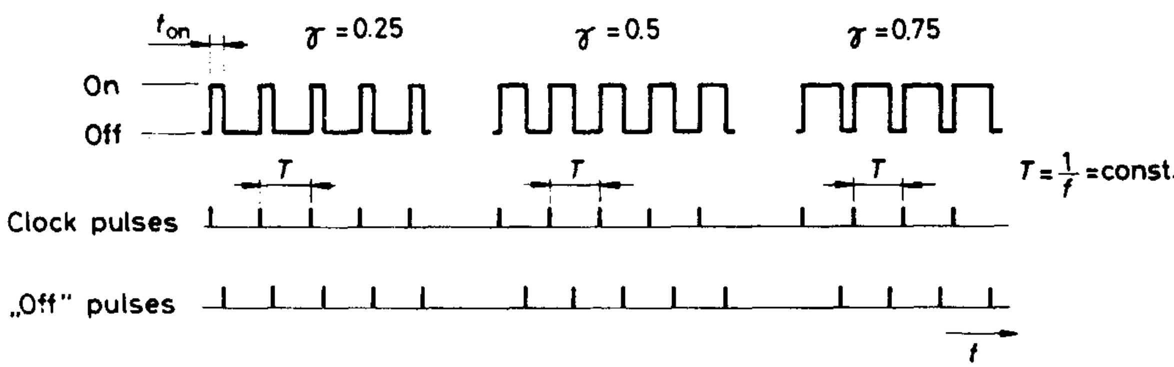

Fig. 2.9. Width-modulated pulse trains of various duty cycles t_{on}/T.

In driving the switching transistor of the energy-transfer circuit, the relative duration of the "on" and "off" times t_{on}/t_{off} can be varied (i.e. the pulse time can be modulated) in one of the following ways:

(1) *Constant frequency* ($f = 1/T$ = constant), variable on and off times (t_{on} and t_{off}) (see Fig. 2.9.);

(2) *Variable frequency*

(a) Constant "off" time t_{off}, variable "on" time t_{on} (see Fig. 2.10. (a)).

(b) Constant "on" time t_{on} variable "off" time t_{off} (see Fig. 2.10. (b)).

(c) Variable "on" time t_{on} and variable "off" time t_{off} (see Fig. 2.10. (c)). This also involves a combination of pulse frequency and width (modulation).

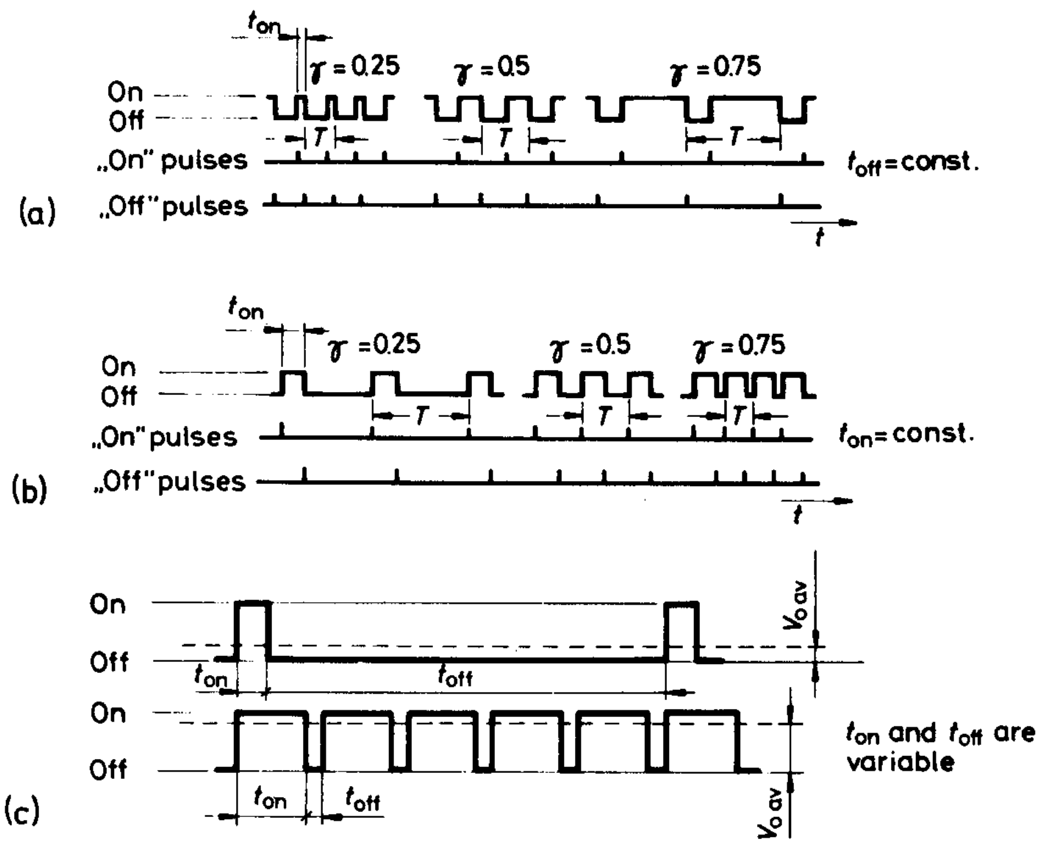

Fig. 2.10. Frequency-modulated pulse trains: (a) t_{off} = constant; (b) t_{on} = constant; (c) t_{on} and t_{off} are varying.

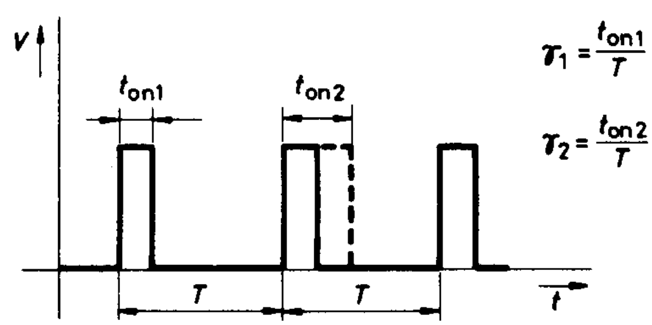

Fig. 2.11. Modulation of the duty cycles of constant-frequency pulses (pulse width modulation).

In the first case, the duty cycle of the pulses is modulated with the instantaneous value of the error signal. The pulse duty cycle is the quotient of the duration and cycle time of the pulses (see Fig. 2.11.):

$$\gamma = \frac{t_{on}}{T} = \frac{t_{on}}{t_{on} + t_{off}}. \tag{2.1}$$

The term "pulse duty-cycle modulation" is also used frequently in the literature. In this case, the power supply proper is referred to as a "PWM switched-mode power supply".

Of the frequency-modulation procedures, the ones described under (a) and (c) are used only rarely, on account of their unfavourable dynamic properties. The latter frequency-modulation circuit (the one described under (c) can be realized by far the simplest circuity. This arrangement is characteristic of the so-called "on-off hysteresis regulation system". Regulators of this kind are referred to as relay or two-position regulators or, derived from the German terminology, as two-point regulators (Zweipunktregelung). On account of their "random" functioning, they have been given the term RANDOM SWITCHING REGULATOR in the English literature.

These two-position regulators perform a continuous comparison of the output voltage and the reference voltage. The bistable circuit element, which may be e.g. a Schmitt trigger, will change state at the instant of the error signal attaining a definite value. The Schmitt trigger will open or close the switching transistor of the energy transfer circuit depending on whether the output voltage is lower or higher than a definite threshold value. Thus, the use of the two-point regulator is

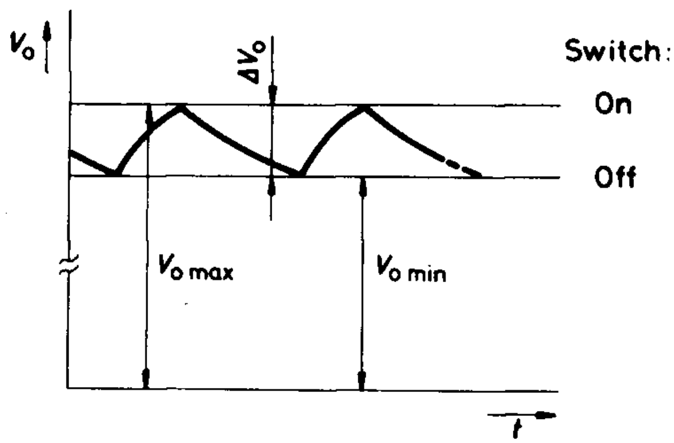

Fig. 2.12. Output voltage dependence of the two-point regulator.

accompanied by periodic variations in the output voltage (see Fig. 2.12.). The voltage variation ΔV_0 can be reduced by increasing the gain.

In pulse-width modulated switched-mode power supplies, the pulses modulated in width act upon the switching transistor at discrete time intervals. Hence, it is often said that the two-point regulators respond to changes in the input voltage and the load current faster than pulse-width modulated regulators. This difference is irrelevant in actual practice. The regulation loop of each switched-mode stabilized power supply also includes a single or multi-component *LC* filter. This filter network, above all, determines the dynamic properties of each switched-mode regulator.

The regulation circuits are discussed in the Section "Regulation and Protection Circuits" (Part III).

2.2.2 PWM Switched-mode Power Supplies

When a constant-frequency system is exposed to varying loads and input voltages, the output voltage is stabilized through variation of the width of pulse passed to the switching transistor (or transistors) of the energy-transfer circuit. The output voltage can be varied from zero to a maximum value by increasing the pulse width from zero to 100 per cent of the switching cycle time (T). The upper limit varies with the output load current and the input voltage.

Figures 2.13. and 2.14. show the block diagrams of a PWM and a push-pull PWM switched-mode power supply, respectively. Both arrangements allow the use of both AC and DC voltages. In either case, the mains voltage is first applied to the high-frequency filter, then two-way rectified by a bridge rectifier and then passed to the energy-storage filter. In this way, the mains voltage of 220 V is converted into an unregulated (crude) DC voltage of about 240 V (with the power-supply output under nominal load). The HF filter reduces the noise fed to the power line, generated by the switching edges of the switching elements.

The DC voltage is now passed to the primary-winding of a ferrite-core transformer (DC–AC inverter) through an adequate switching stage. In the second diagram the DC voltage is passed to the switching transistor which works in push-pull mode. In either case, the voltage appearing at the secondary winding of the transformer is rectified by fast switching diodes and filtered.

The comparator and error-signal amplifier circuit comprises several active elements. This circuit compares the reference voltage with the pilot voltage V_{pil}

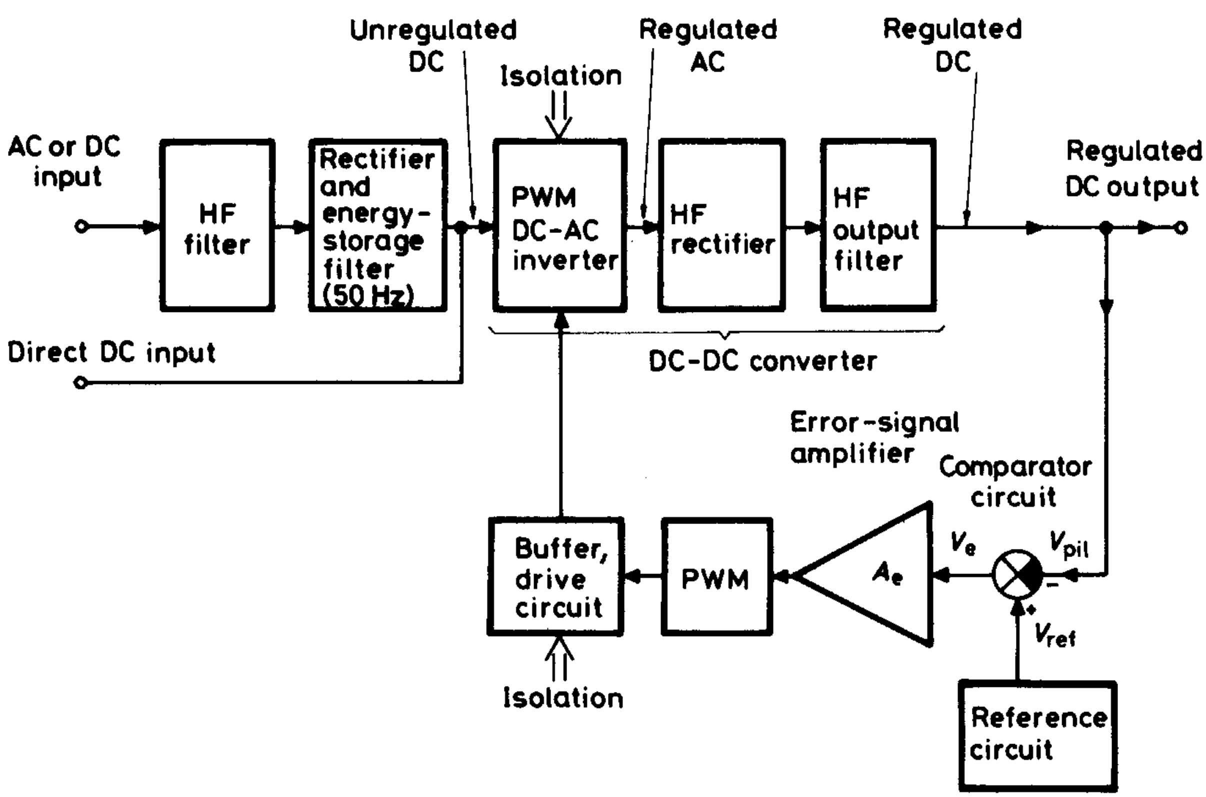

Fig. 2.13. Block diagram of pulse-width-modulated (PWM) switched-mode power supply.

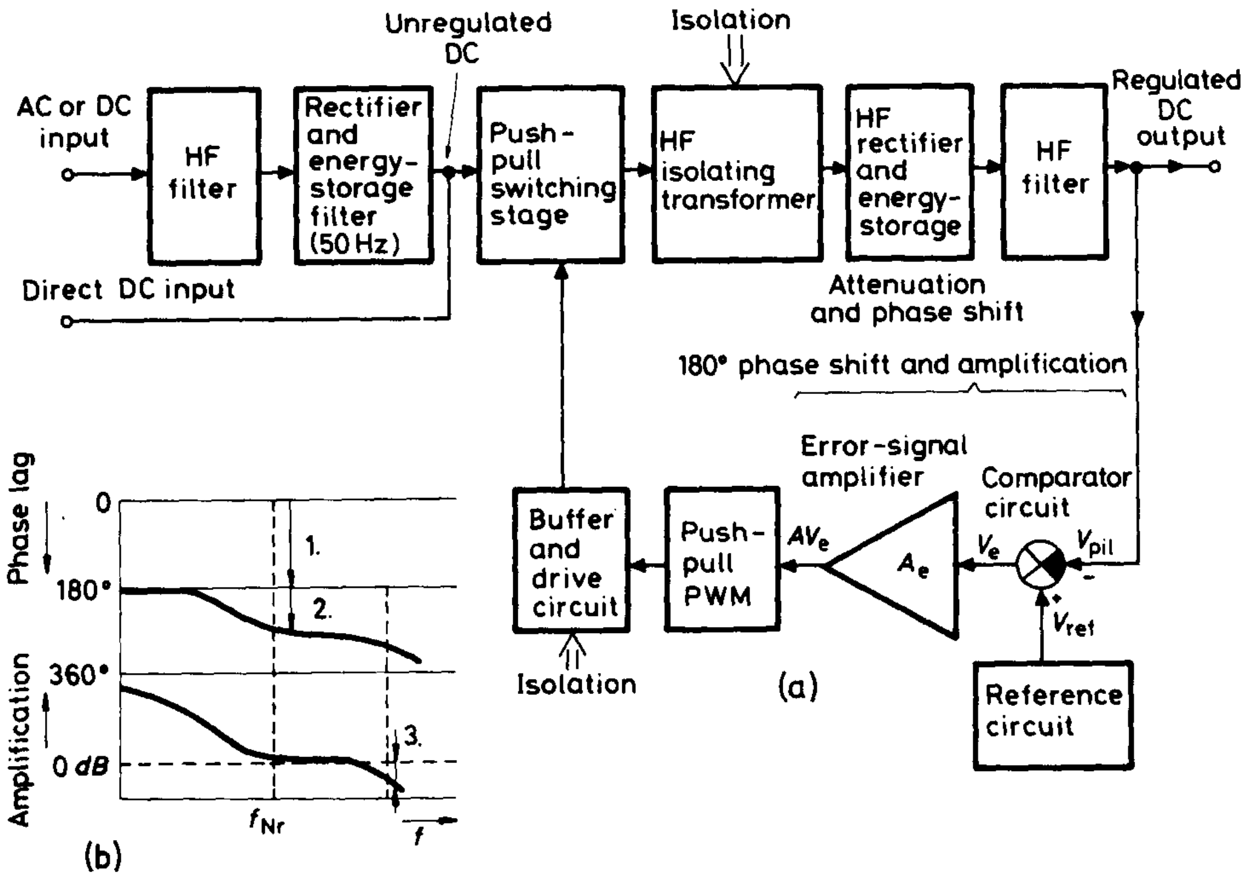

Fig. 2.14. Push-pull pulse-width modulated (PWM) switched-mode power supply: (a) block diagram; (b) typical frequency response of the loop (*1* phase lag on account of the negative feedback; *2* phase lag on account of the output filter and the error-signal amplifier; *3* the gain is already less than unity before the phase shift of 360°; f_{Nr} = nominal resonance frequency).

taken from the output and amplifies the error signal to a suitable value to drive the pulse-width modulator. The push-pull PWM applies the width-modulated alternately to one or the other transistor in the energy transfer circuit. Isolation of the output of the power supply from the input is made possible by the transformer in the energy-transfer circuit and by the base-driving transformer of the switching transistor . The transformer-type base drive ensures a DC coupling between the driving and the feedback circuits and the output. In this way, p roblems associated with the AC coupling of the feedback circuit can be avoided.

PWM switched-mode power supplies have a constant switching frequency. The advantage of the fixed frequency is that it cannot drop into the audible range, as may sometimes occur with variable-frequency systems under actual load conditions. Another advantage of fixed frequency operation is that, under normal conditions of operation, external filters may also be added to the unit so that a maximum attenuation is obtained at a particular frequency. They cannot be used over such broad ranges of input and output voltages as frequency-modulated systems; furthermore, some versions of this circuit arrangement are apt to jump into uncontrolled mode of operation (see later).

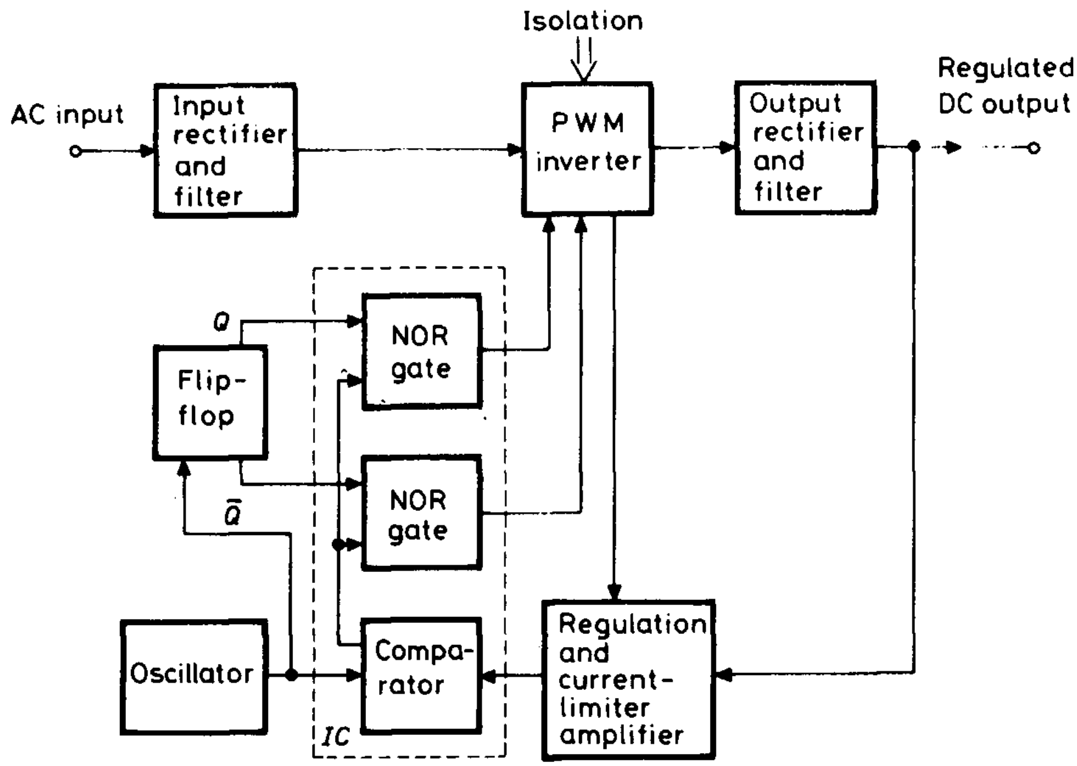

Fig. 2.15. Block diagram of switched-mode power supplies manufactured by Lambda.

The figures below show the block diagrams of switched-mode power supplies manufactured by Lambda and Powertec, respectively.

Figure 2.15. shows the simplified block diagram of the Lambda switched-mode power supplies. In this circuit arrangement, the push-pull driving signals are provided by two NOR gates. Together with the two NOR gates, the comparator is accommodated in a single integrated circuit. The operation of this regulated circuit is discussed at length in the Section "Regulation and Protection Circuits".

Figure 2.16. shows the block diagram of the 9L5-120 Powertec switched-mode power supplies. The power supply has an output isolated from the mains and a slow turn-on circuit. The actual power supply of 5 V output voltage is capable of providing a load current of 120 A at an efficiency of 80 per cent. The power loss in the power supply is only 120 W, representing a reduction of power dissipation to one tenth of that of linear series-regulation power supplies of similar powers. This is because the efficiency of conventional linear series-regulation power supplies of 5 V output voltage is about 33 per cent, representing a power loss of 1200 W at a load current of 120 A.

As can be seen from the block diagram, the output voltage is passed to the slow turn-on and rectifier circuits via the input fuse. This is necessary because the current surge at the instant of turn-on may lead to the destruction of the switching transistor and the rectifier components (the output capacitor being uncharged). The slow turn-on circuit of the Powertec design (also referred to as a soft start circuit) includes a thyristor bridge-circuit rectifier (two thyristors and two power diodes). When the power supply is turned on, the two thyristors increase the output power to the ultimate value with a slow phase regulation. This process retards the turn-on current surge, thus protecting the circuit elements of the input stage. After the initial turn-on period, the thyristors will act as conventional rectifiers, not introducing any appreciable power loss.

The unit operates at a fixed frequency of 20 kHz. The "control circuit" is the proper pulse-width modulator. As can be seen, the output of 5 V d.c. is isolated

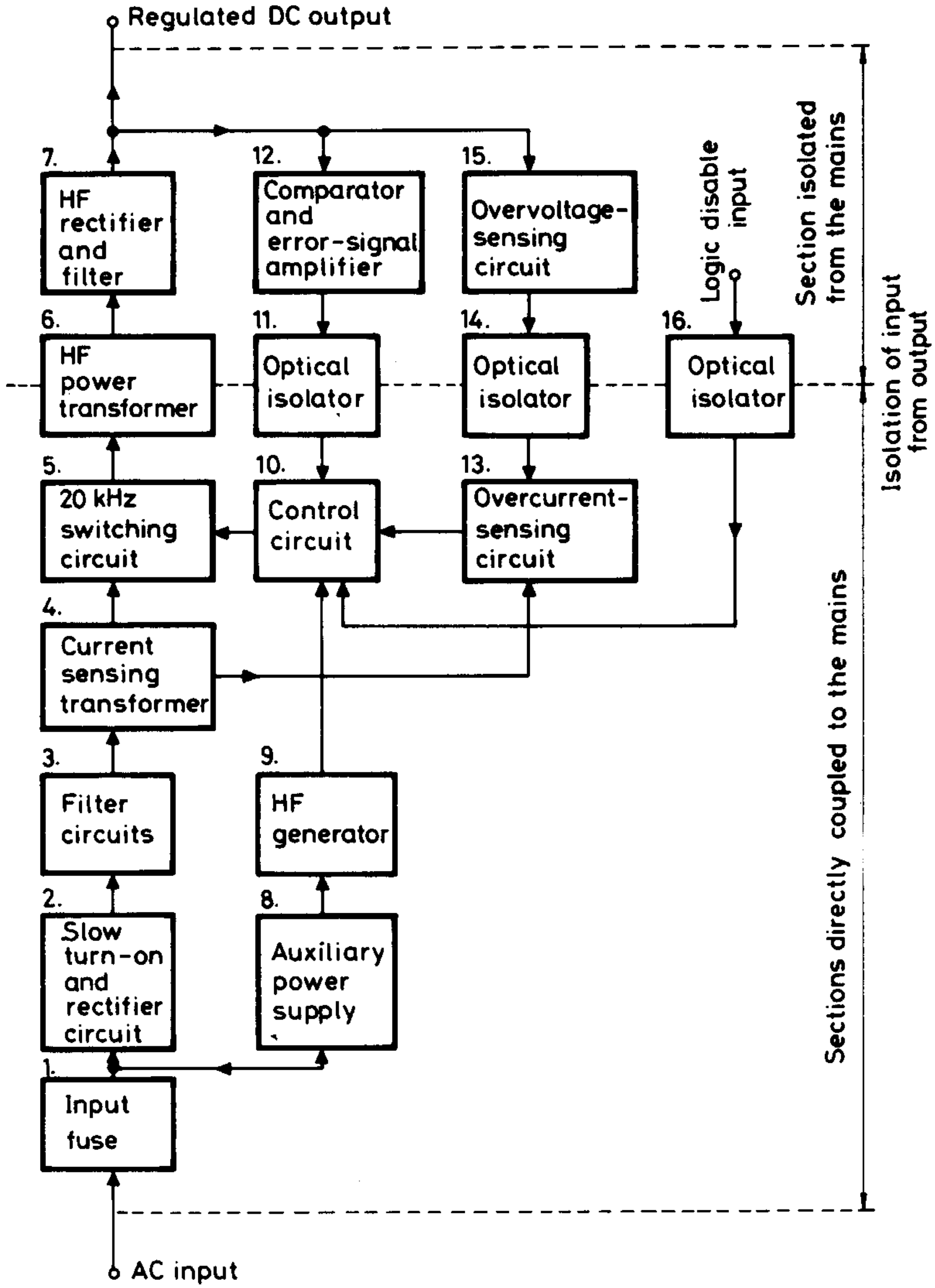

Fig. 2.16. High-efficiency switched-mode stabilized power supply with output isolated from the mains, employing a slow turn-on circuit (Powertec).

completely from the input by a power transformer and three optical isolators (see the unit along the vertical dashed line). One circuit innovation in this power supply is the use of optical isolators (also termed optical couplers) with linear response characteristics.

The pulse-width modulated pulses of 20 kHz coming from the transistor switching circuit are passed to a HF power transformer; afterwards they are rectified and filtered. Though appearing to be a simple procedure, this requires the use of special rectifier diodes in practice. Commercially available fast silicon power diodes exhibit a voltage drop of about 1.3 V at a current of 120 A (156 W), so they are unsuitable for use as the rectifiers of low-voltage high-current power supplies.

The Powertec designers use special Schottky (hot-carrier) power diodes for the 20 kHz rectifiers. These exhibit a voltage drop of only 0.4 V at 120 A, representing a power dissipation as low as 48 W. As a result, these diodes will increase the efficiency automatically. The efficiency of 80 per cent mentioned above would have been impossible without the use of the Schottky power diodes.

Typical data for mass production power supplies are $V_o = 5$ V ± 5 per cent; $I_{o\max} = 120$ A; $S_i \leqq \pm 0.1$ per cent; $S_L \leqq \pm 0.1$ per cent; $u_{rpp} \leqq 50$ mV. The overcurrent limiter is activated at 125 per cent of $I_{o\max}$. The overvoltage protection circuit is tripped at 6.5 V ± 5 per cent. The maximum operating temperature is 40 °C.

2.2.3 Frequency-modulated (PFM) Switched-mode Power Supplies

Of the frequency-modulation procedures, the fixed pulse-width version (i.e. the one applying a constant "on" time, $t_{on} =$ constant) is by far the most commonly used.

If the "on"' time t_{on} is taken as constant, the automatic circuit will maintain the output voltage level constant under conditions of varying load current or input voltage by varying the frequency, i.e. the number of pulses in unit time.

Figure 2.17. shows the block diagram of a PFM switched-mode power supply. Compared with the previous block diagram, the only difference here is that the error-signal amplifier drives a variable-frequency pulse generator (i.e. a frequency modulator).

Frequency modulation with a fixed "on" time provides a system free of restrictions; thus it is much more capable of dealing with troubles due to an instantaneous overload or drop in the mains voltage. Since the pulse width determines the ripple voltage for a given inductance and capacitance, the ripple will remain constant with a change in the mains voltage and the load. The circuit remains operative even under very wide variations in input voltage. The output voltage is variable over broad limits. The principal drawback is that it may penetrate the audio-frequency range at low loads. Furthermore, since the ripple frequency is variable, it is less easy to filter out. To avoid penetrating the audio-frequency range, with either power supply, the simplest procedure is to use a pre-loading resistor to set the minimum frequency with the power supply in the "idle"

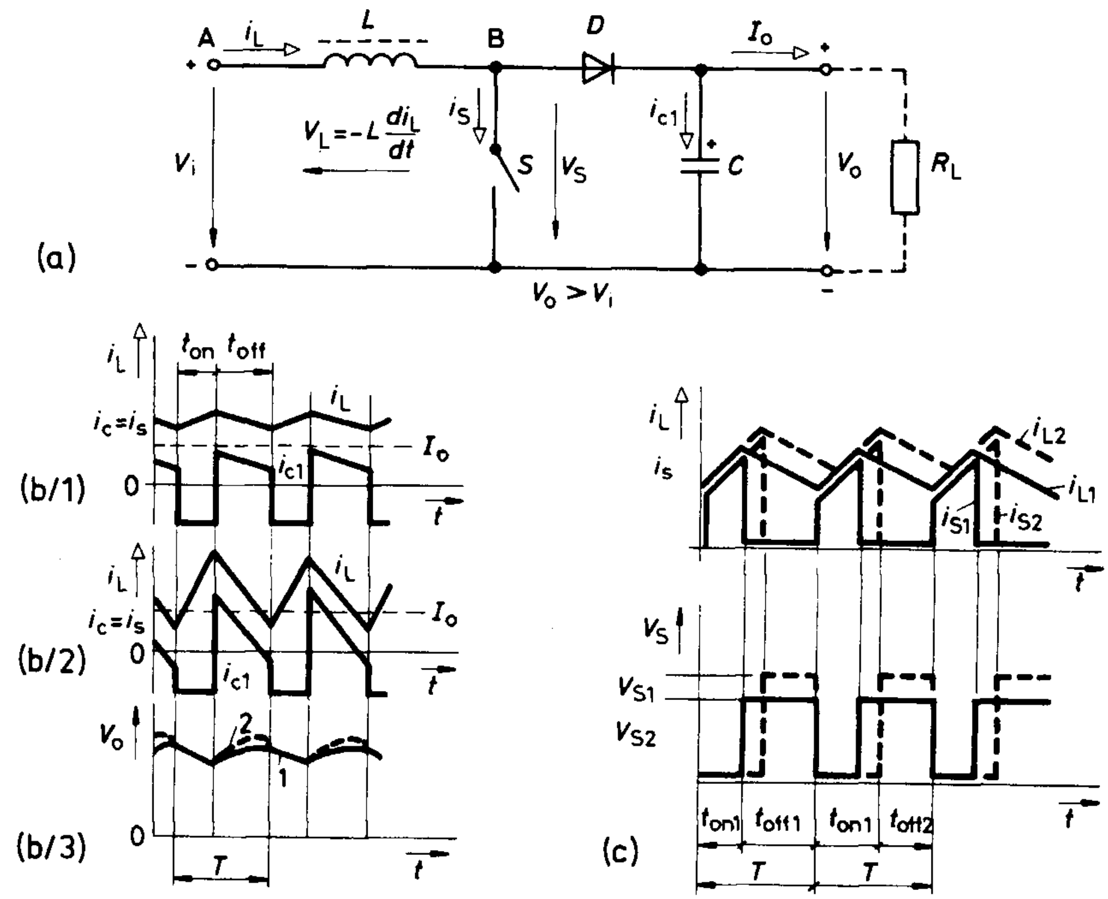

Fig. 3.18. Basic voltage-step-up energy-transfer circuit employing a choke (a) and typical waveforms (b) and (c).

obtained. The terminal lead of the output sensor must be connected to the 300 V terminal.

The overall circuitry of DC step-down switched regulators will be described in Part V, under the heading "Practical Circuit Arrangements".

3.2 DC Step-up Circuits

3.2.1 Basic Circuits and their Operation

If a stable voltage higher than the input voltage is required, a DC step-up energy transfer circuit is employed in the switched-mode regulator.

Figure 3.18. shows the basic circuitry with the appropriate time graph.

In this energy-transfer circuit, the energy is transferred to the output with the switch in the "off" state (the transistor being cut off). Hence, it is also referred to as a flyback energy-transfer circuit or converter.

When the switch is turned on, the current through inductance L rises linearly from the lowest value prevailing at the instant of turn-on ($I_{L_{min}}$) to the maximum value attained at the instant of turn-off ($I_{L\,max}$). The diode is required to prevent the storage capacitor (C) from being discharged during the "on" period of the switch. During this period of time the energy demand of the load is satisfied by storage capacitor C. After the switch is turned off, the current through choke L begins to decline, decreasing from $I_{L\,max}$ to $I_{L\,min}$ during the length of time t_{off} (Figs. 3.18. (b) and (c)). Meanwhile, the inductance transfers the energy accumulated within it to the load. Here the underlying process is that the voltage induced in the choke is added to the input voltage at the instant of the switch being turned off. Put in another way, an energy $W=0.5\,Li^2$ stored in the choke produces a voltage at point B which is higher than that at point A. The output voltage can be regulated by varying the "on" time of the switch. A longer "on" time results in a higher output voltage (see Fig. 3.18. (c)).

With switch S in the "on" state, the entire input voltage is passed to inductance L. Thus,

$$V_i = L\frac{di_L}{dt}. \tag{3.38}$$

Let us assume the input voltage V_i to remain constant during the t_{on} time. Integrating eqn. (3.38),

$$i_L = I_{L\,min} + \frac{V_i}{L}t. \tag{3.39}$$

Substituting the value of $t=t_{on}$ in the above expression, an equation is obtained that establishes a relationship between the lowest and highest values of the inductance current:

$$I_{L\,max} = I_{L\,min} + \frac{V_i}{L}t_{on}. \tag{3.40}$$

When the switch is turned off, the current flowing in the inductance now flows through the load resistor R_L, the inductance and the power source. It may be written that

$$V_i - V_o = L\frac{di_L}{dt}. \tag{3.41}$$

Integrating this equation,

$$i_L = \frac{V_i - V_0}{L}t + I_{L\,max}. \tag{3.42}$$

After the elapse of time t_{off}, the current flow in the inductance is

$$I_{L\,min} = \frac{V_i - V_o}{L}t_{off} + I_{L\,max}. \tag{3.43}$$

Eliminating the difference $I_{L\max} - I_{L\min}$ from eqs (3.40) and (3.43),

$$V_o = V_i\left(1 + \frac{t_{on}}{t_{off}}\right) = \frac{V_i}{1-\gamma}. \tag{3.44}$$

It can be concluded from this expression that whatever the value of the pulse duty cycles ($\gamma \neq 0$), the output voltage will invariably be greater than the input voltage, i.e. $V_o > V_i$.

The voltage V_s across the terminals of the switch is equal to the output voltage V_o provided that the voltage drop across diode D is ignored:

$$V_s = V_i\left(1 + \frac{t_{on}}{t_{off}}\right) = \frac{V_i}{1-\gamma}. \tag{3.45}$$

The maximum current flowing through the switch ($I_{s\max}$) is equal to $I_{L\max}$. Thus the mean value of current flowing through the switch is

$$I_{sav} = \frac{I_{L\min} + I_{L\max}}{2}\gamma = \frac{I_{L\min} + I_{L\max}}{2}\frac{t_{on}}{T}. \tag{3.46}$$

Since

$$I_o = \frac{I_{L\min} + I_{L\max}}{2}\frac{t_{off}}{T}, \tag{3.47}$$

then

$$I_{sav} = I_o \frac{t_{on}}{t_{off}}. \tag{3.48}$$

To allow calculation of the current $I_{s\max} = I_{L\max}$, eqs (3.40), (3.47) and (3.43), (3.47) are combined. By eliminating $I_{L\min}$, the expressions

$$I_{s\max} = I_o \frac{T}{t_{off}} + \frac{V_i}{2L} t_{on} \quad \text{and} \tag{3.49}$$

$$I_{s\max} = I_o \frac{T}{t_{off}} - \frac{V_i - V_o}{2L} t_{off} \tag{3.50}$$

are obtained for the maximum switch current $I_{s\max}$.

The operational requirements of the switch are given in broad outline by eqs (3.45), (3.48), (3.49) and (3.50). These enable the type of the switching element to be selected in advance.

At the instant of the switch being turned off, the capacitor begins to be charged (see Fig. 3.18. (b)), so the output voltage rises. During this time interval, the charging current of the capacitor (i_c) varies in accordance with the equation

$$i_c = i_L - I_o, \tag{3.51}$$

where i_L is the time function to be obtained from eqn. (3.42).

If $I_o < i_L$ during the entire "off" period of the switch, the voltage will rise monotonously at capacitor C. The output voltage attains its maximum level at the instant of the switch being turned on; afterwards it begins to fall. This process is repeated in a cyclic manner. In this case, the double amplitude of the AC voltage component may be taken as equal to the voltage drop across the capacitor after the elapse of "on" period t_{on}:

$$\Delta V_{\sim} = \frac{1}{C}\int_{0}^{t_{on}} i_c \, dt . \tag{3.52}$$

Assuming the load current to be nearly constant,

$$\Delta V_{\sim} = \frac{I_o t_{on}}{C} . \tag{3.53}$$

Apparently, the AC component of the output voltage depends wholly on the capacitance. However, this statement is only valid if $I_o < I_{L\,min}$, i.e. if the choke has a sufficiently high inductance. For a low inductance value, current i_L falls rapidly, and $I_{L\,min}$ may be lower than load current I_o. In this case, the capacitor begins to be discharged before the switch is turned on. As a result, however, the AC voltage component increases. The characteristics of currents i_L, i_c and output voltage V_o are shown in Fig. 3.18. (b) 1 for a high inductance L; Fig. 3.18. (b) 2 shows the same characterisstics for a low inductance.

The minimum value of inductance L, at which the AC component of the output voltage becomes dependent on capacitance C can be obtained from the condition $I_o <_{L\,min}$. The calculated inductance L must satisfy the inequality

$$L > \frac{t_{off}^2 (V_o - V_i)}{t_{on} I_o} . \tag{3.54}$$

To sum up, the following principal relationships may be used in the design of the circuit (see also the design of the circuit in Fig. 3.3.).

When the inductance is never free of energy ($L \geqq L_{min}$), the output voltage is

$$V_o = \frac{V_i}{1-\gamma}, \text{ if } I_o > I_{lim} , \tag{3.55}$$

where I_{lim} is the limit current.

When the inductance is free of energy during part of the cycle, the output voltage is

$$V_o = V_i + \frac{V_i^2 \gamma^2 T}{2LI_o}, \text{ if}$$

$$0 \leqq I_o \leqq I_{lim} , \tag{3.56}$$

where I_{lim} is the limit current
T is the cycle time.

If the series loss resistances are also taken into account (Fig. 3.4.) the output voltage (at a load above limit current $I_{\lim}$) will be

$$V_o = \frac{V_i}{1-\gamma} - I_o \frac{\gamma R_{sat} + (1-\gamma) R_D + R_{LS}}{(1-\gamma)^2}. \tag{3.57}$$

The limit current is

$$I_{\lim} = \frac{\gamma T V_i}{2(1-\gamma)L}. \tag{3.58}$$

The peak voltage and peak current at the transistor and the diode are equal to the output and the $I_o/(1-\gamma)$ values, respectively.

The efficiency of the circuit is

$$\eta = \frac{P_o}{P_i} = 1 - \frac{I_o}{(1-\gamma) V_i} [\gamma R_{sat} + (1-\gamma) R_D + R_{LS}]. \tag{3.59}$$

Capacitive Step-up Energy-transfer Circuit

The circuit shown in Fig. 3.19. may be employed when a low output power is required. It contains two switching elements that are turning on and off alternately. Let us assume both switching elements to be in the "off" state. Now capacitor C_2 is charged to the input voltage through diodes D_2 and D_3. When switch S_B is turned

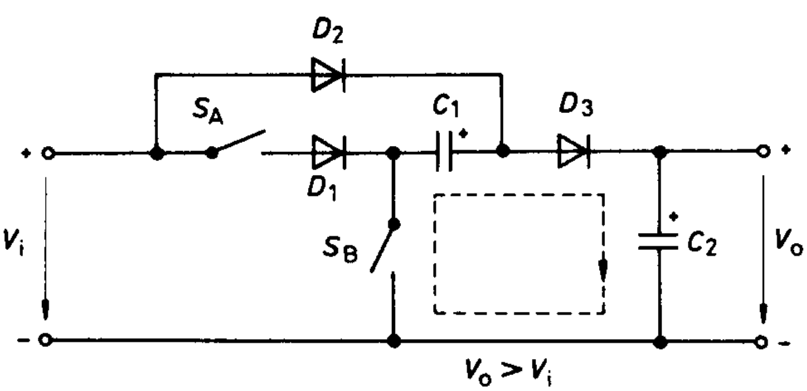

Fig. 3.19. Low-power capacitive voltage-stepup energy-transfer circuit.

on, capacitor C_1 is also charged through diode D_2. When switch S_B is turned off, S_A is turned on. Thus, the output storage capacitor C_2 receives the sum of the voltage at capacitor C_1 and the input voltage through diode D_3.

Integrated circuits capable of a push-pull operation may be used for regulation circuits (e.g. SG 1524, MC 3420, ZN 1066E, TDA 1640, TDA 1641, etc.).

3.2.2 Versions of Basic Circuits

The basic step-up circuit may contain a three-terminal choke to reduce the voltage reaching the terminal lead of the winding (i.e. the current flowing through the turns; in this way the operational conditions of the switching element can be alleviated. The underlying principles of operation for the circuits shown in Fig. 3.20. are identical with those described for the basic circuitry (see the discussion of Fig. 3.18.). This time, however, different expressions are obtained for the principal quantities determining the conditions of operation.

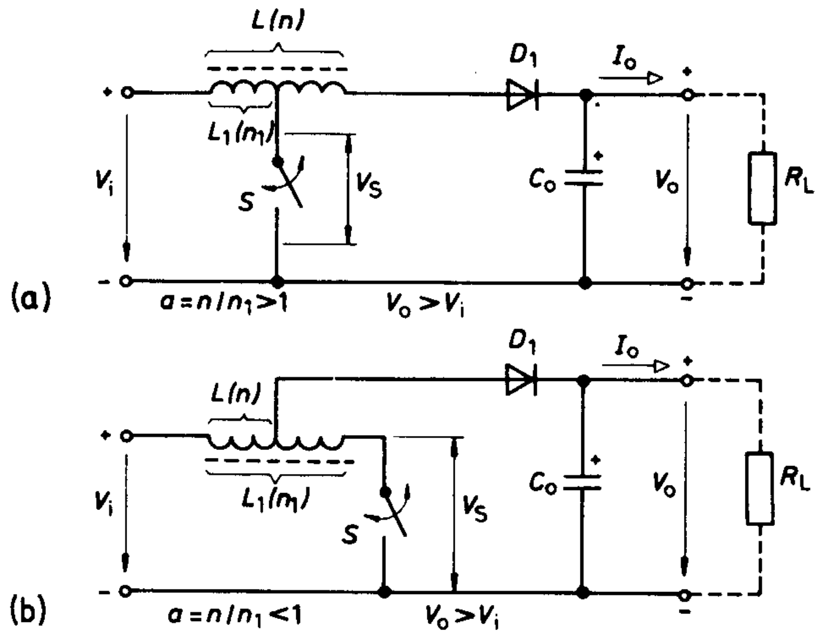

Fig. 3.20. Versions of basic DC voltage-step-up energy-transfer circuit.

Designing Considerations

A detailed analysis of these circuits is given in References [3.1] and [3.7]. It will suffice here to give the final results.

At the instant of the switch being turned off, the maximum current in a coil of inductance L (n number of turns) is

$$I_{\mathrm{L\,max}} = I_{\mathrm{s\,max}}/a\,, \tag{3.60}$$

for each circuit of Fig. 3.20., where

$$a = n/n_1 = \sqrt{L/L_1}\,.$$

After the switch has become conductive, the current flowing through a coil of inductance L begins to rise again from

$$I_{\mathrm{L\,min}} = I_{\mathrm{s\,min}}/a\,. \tag{3.61}$$

The output voltage of the energy transfer circuit may be obtained from the expression

$$V_o = V_i\left(1 + a\frac{t_{on}}{t_{off}}\right). \tag{3.62}$$

At any actual value of a and t_{on}/t_{off}, the output voltage is invariably higher than the input voltage. When $a = 1$, the above expression becomes equal to eqn. (3.44).

The voltage appearing at the terminals of the open switch is

$$V_s = V_i\left(1 + \frac{t_{on}}{t_{off}}\right) = \frac{V_i}{1-\gamma}. \tag{3.63}$$

It is evident from this expression that the voltage across the switching element is unaffected by the characteristics of the inductance, and is equal to eqn. (3.44) for the case of $a = 1$.

The average and the maximum values of the switched current are given by the expressions

$$I_{s\,av} = aI_o\frac{t_{on}}{t_{off}}; \tag{3.64}$$

and

$$I_{s\,max} = aI_o\frac{T}{t_{off}} + \frac{V_i}{2L_1}t_{on}, \tag{3.65}$$

respectively.

Equation (3.53) continues to be valid in the calculation of the output AC voltage component, providing eqn. (3.54) is also satisfied.

The circuits of Figs 3.18. and 3.20. have the following output power:

$$P_o = V_o I_o, \tag{3.66}$$

where I_o is the load current.

This power can be broken up into two parts for these DC step-up energy transfer circuits. One part (P_{o1}) is passed directly from the input to the load at the output terminals, without any intermediate modification. The other part (P_L) comes from the energy accumulated in the inductance. Neglecting the losses due to the circuit components,

$$P_o = P_{o1} + P_L. \tag{3.67}$$

The switching element will vary the inductive power component P_L. Since P_{o1} and P_L are proportional to V_i and

$$L\frac{di_L}{dt},$$

respectively,

$$P_{o1} = V_i I_o, \tag{3.68}$$

$$P_L = (V_o - V_i)I_o. \tag{3.69}$$

A reduction of the inductive power component P_L will allow use of smaller coils.

During the "off" period the current through the inductance falls from $I_{L\,max}$ to $I_{L\,min}$. Meanwhile, part of the inductive energy accumulated in it is passed to the output. Hence the inductive power component P_L may also be written as

$$P_L = \left(\frac{1}{2} L I_{L\,max}^2 - \frac{1}{2} L I_{L\,min}^2\right)\frac{1}{T} =$$

$$= \frac{1}{2} L(I_{L\,max}^2 - I_{L\,min}^2)\frac{1}{T}. \tag{3.70}$$

When the switch is turned on, the energy within the inductance attains the initial value again after t_{on} time, as the current in inductance L rises from $I_{s\,min}$ to $I_{s\,max}$. Thus

$$P_L = \left(\frac{1}{2} L_1 I_{s\,max}^2 - \frac{1}{2} L_1 I_{s\,min}^2\right)\frac{1}{T} =$$

$$= \frac{1}{2} L_1 (I_{s\,max}^2 - I_{s\,min}^2)\frac{1}{T}. \tag{3.71}$$

The two last equations show how the inductive power component P_L affects the current flowing through inductances L_1 and L, and through the switching element.

To better illustrate the influence of factor a, Fig. 3.21. shows the set of curves of function $V_i/V_o = f(a)$ plotted for different fixed values of the pulse duty cycle.

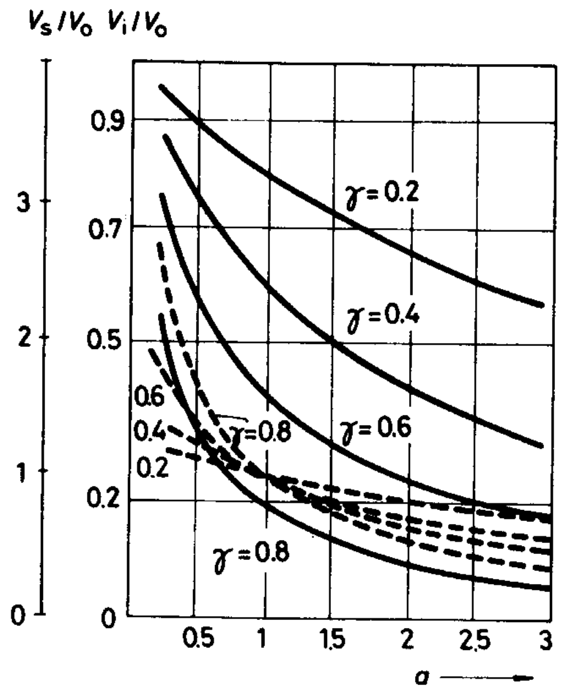

Fig. 3.21. Graphs for designing the basic DC voltage-stepup energy-transfer circuit. The curves drawn in continuous and dashed lines represent the functions $V_i/V_o = f(a)$ and $V_{sw}/V_o = f(a)$, respectively. The pulse duty cycle γ is the variable.

Using eqn. (3.62), the function $V_i/V_o = f(a)$ may be written

$$\frac{V_i}{V_o} = \frac{1-\gamma}{1+\gamma(a-1)}. \tag{3.72}$$

It is evident from the curves shown in solid line that, with an increased value of a, changes in the duty cycle γ have a diminishing effect on the output voltage. This means a narrowing of the voltage regulation range. This is why it is inadvisable to use a three-terminal choke that has a high a value.

At high a values, the output voltage may be several times as high as the input voltage. This, in turn, means that power component P_L greatly exceeds P_{o1}.

Figure 3.21 also shows, as dashed lines, the set of curves $V_s/V_o = f(a)$. Again, the parameter is pulse duty cycle γ. The mathematical form of this function can be obtained using eqs (3.62) and (3.63), by elimination of the quantity V_i:

$$\frac{V_s}{V_o} = \frac{1}{1+\gamma(a-1)}. \tag{3.73}$$

It is apparent from the dashed curve that, with the switch turned off, the voltage V_s appearing at the terminals of the switch may exceed the output voltage.

As with the choke with two terminal leads, the calculated inductance L of the three-terminal choke must also satisfy eqn. (3.54). Since the rate of increase in the current flowing through the switch in the "on" state depends on inductance L_1, the latter must be selected so that current $I_{s\,max}$ cannot exceed the maximum permissible collector current $I_{C\,max}$ even under the worst conditions. Accordingly, the inequality

$$L_1 > \frac{V_i t_{on}}{\left(I_{C\,max} - I_o a \cdot \dfrac{T}{t_{off}}\right)^2} \tag{3.74}$$

can be deduced easily by the use of eqn. (3.65).

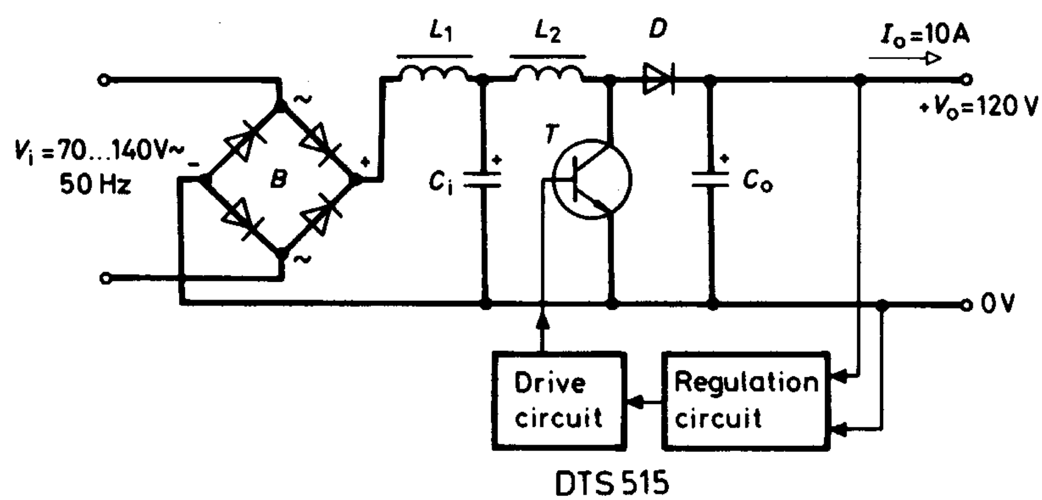

Fig. 3.22. "Step-up" switched-mode regulator (DTS 515 of Delco Electronics).

3.2.3 Examples of Circuitry

Figure 3.22 shows the schematic circuit arrangement of a "step-up" switched-mode regulator. Providing an output power of 1200 W, this DC stepup energy transfer circuit has been realized with a single switching transistor type DTS 515. The principal parameters of this transistor which was developed specifically for switched-mode operation are discussed in connection with the components.

Figure 3.23 shows a DC stepup switched-mode regulator with 100 W output power. The energy-transfer circuit employs a DTS 1020 transistor in a Darlington configuration.

Each of the resistors in the circuit has a power rating of 0.5 W.

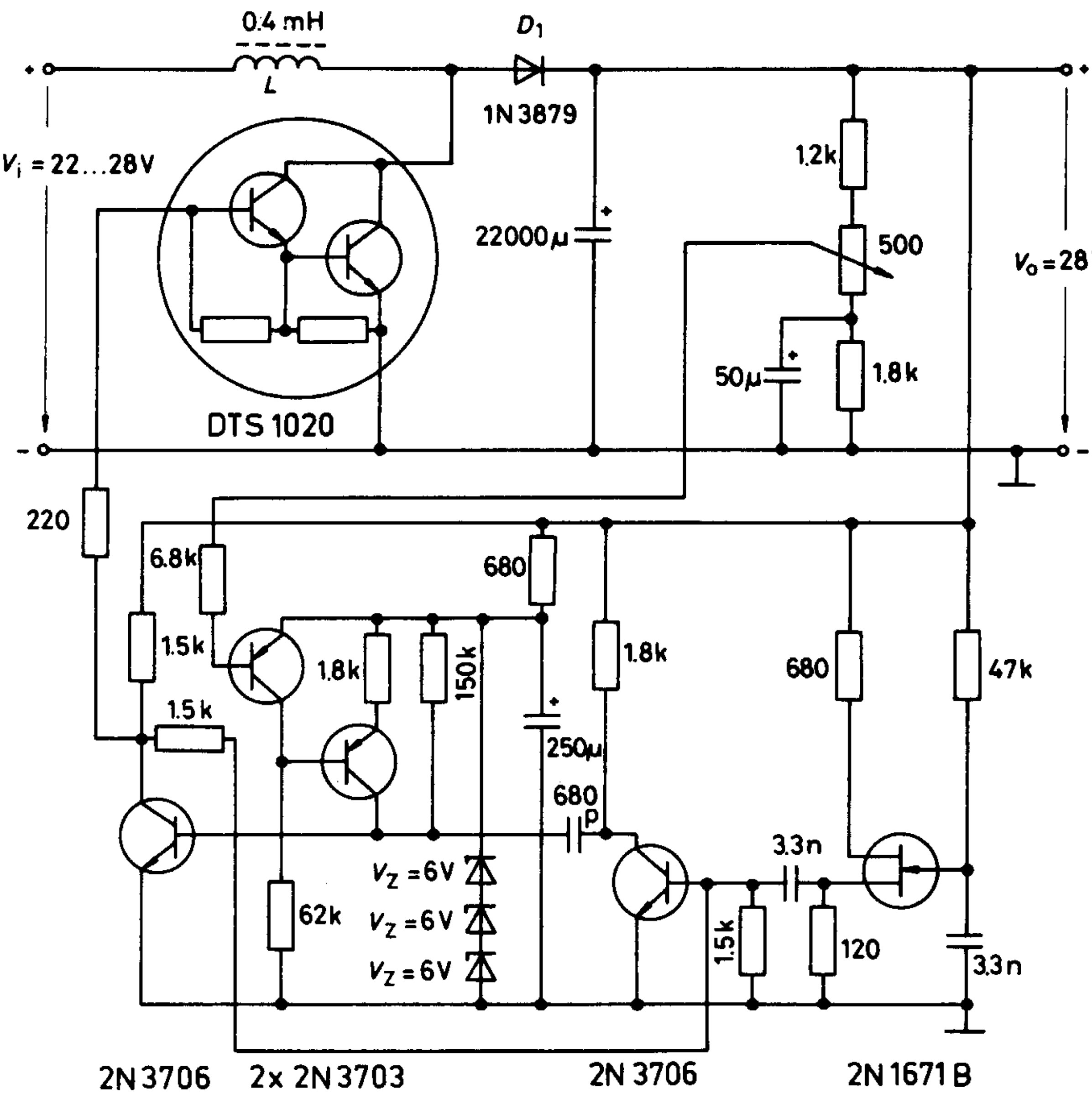

Fig. 3.23. DC voltage step-up switched-mode regulator with 100 W output power (Delco Electronics).

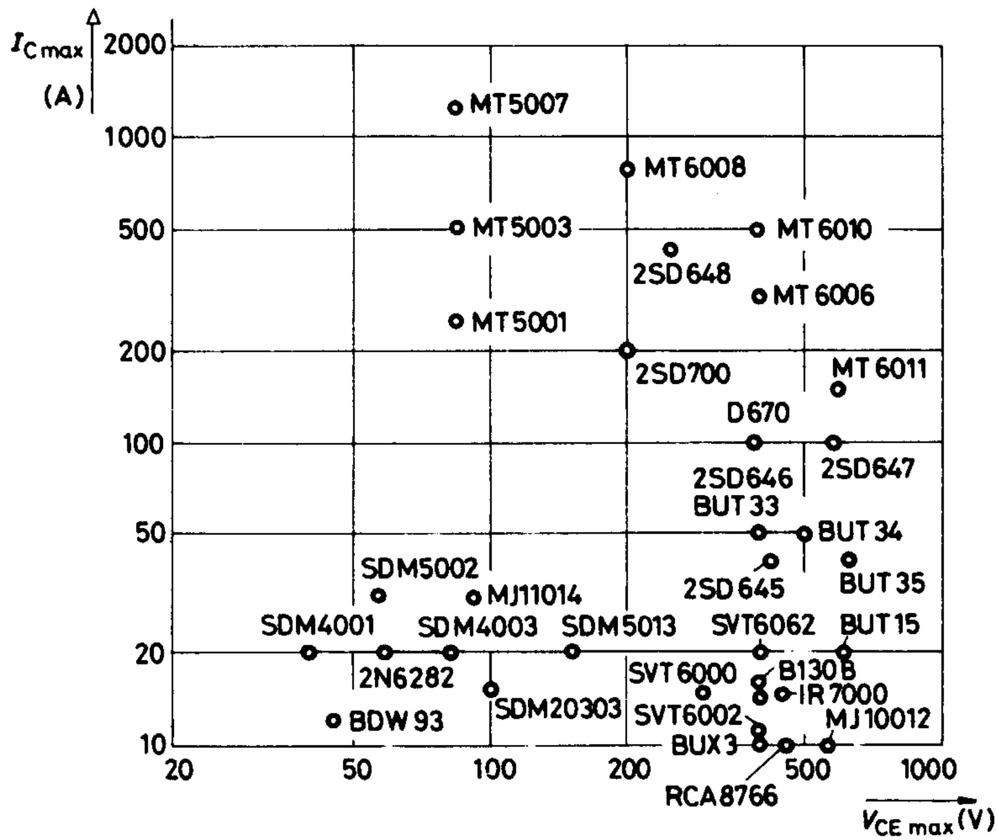

Fig. 6.10. Very high-power Darlington transistors.

— No secondary breakdown. There is no need for a protection circuit. Simple RC circuits may be used.

— High-speed switching rate.

As an example, the Siliconix VN 64 GA has average rise and fall times of about 45 ns when driven from a source impedance of 50 ohms. At such short rise and fall times minimum switching losses are incurred. Thus the operating frequency may be as high as 500 kHz, largely reducing the size and cost of the filter components.

— Lack of minority storage time.

The turn-off storage delay time may have ruinous effects on push-pull converters employing bipolar transistors, because of the hazards of the two transistors passing current simultaneously. These include high peak currents, extreme dissipation, saturation of the core and finally the destruction of the device. To eliminate this, converters employing bipolar transistors have to be provided with circuits preventing the saturation of the core.

— No current limiting (hogging).

To obtain higher output powers, the VMOS power FET's can be connected directly in parallel with each other, without the necessity of compensation circuits.

— Very high gain (10^6–10^7).

— Low power loss. With the transistors in parallel connection, the power loss for the "regulation component" is reduced to a very low value, even at high currents.

Although VMOS power FET's appear to be ideally suitable for replacing the bipolar switching transistors, they have several drawbacks. However, these drawbacks can be eliminated by careful design. Some of them are listed below.

— High "on" resistance.

The saturation loss is relatively high for devices available at competitive prices. As for the future, a considerable decline in the saturation losses of VMOS devices is expected, because the new ones will have much lower "on" resistances than the cheaper currently mass-produced types. As an example, of the 9.7 W power loss of a power supply with 10 A loadability the forward loss of the rectifier diodes is 5.2 W, whereas that of a VMOS power FET saturation loss is only 2.7 W. The remaining losses are mainly switching losses made up of ferrite and copper losses in the transformer and the choke, and the losses in the regulation circuits. In the case of a low-output-voltage power supply, the pass (forward) losses of the rectifier diodes represent the most significant loss mechanism. In current low-prices high-performance power supplies, the power FET's are employed in the driving circuits in front of the power switching stage.

— Extremely high gain (10^6–10^7).

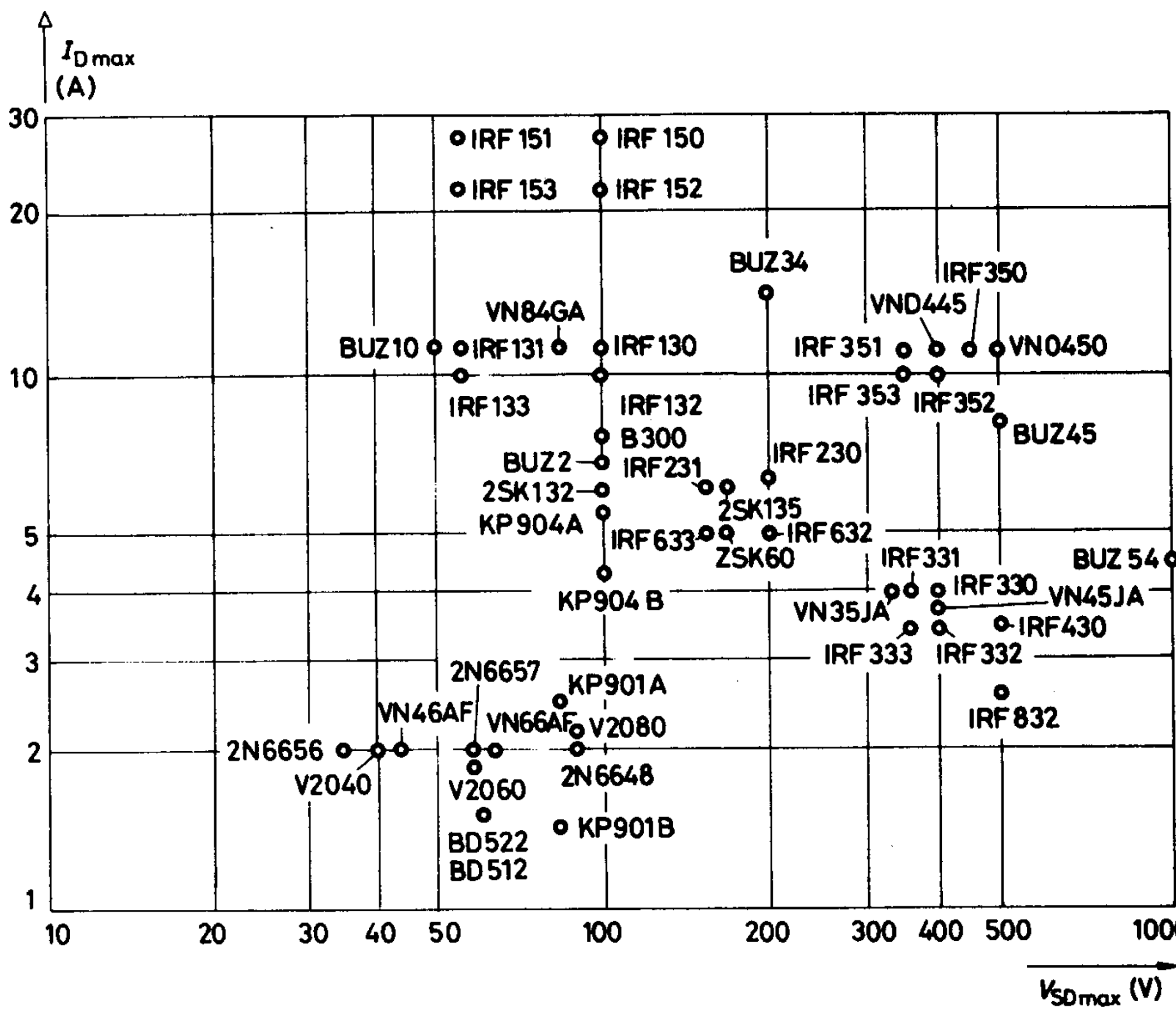

Fig. 6.11. Power MOS FET's.

This parameter may give rise to parasitic oscillations of very high frequency—even when the VMOS FET is working alone. To suppress the parasitic oscillation, a capacitor of 82 to 100 pF should be connected across the drain and the source.

— Low breakdown voltage.

The less expensive devices have a relatively low breakdown voltage (90 V). In applications requiring a higher breakdown voltage (e.g. 350 or 400 V), the use of Siliconix VN 35 JA or VN 45 JA devices offers a possible but expensive alternative.

— Long lead times.

This is a drawback; in short-circuit mode the "lead" times may become even longer. Together with current limiting, this problem has to be considered when a power supply containing a regulation circuit is to be designed.

Figure 6.11. shows a few types of MOS power switching FET's Reference [6.9]. In the diagram, the maximum drain-source voltage ($V_{SD\,max}$) and the maximum drain current ($I_{D\,max}$) are represented by co-ordinates x and y, respectively.

Several different technologies are applied in the manufacture of MOS power FET's: VMOS power FET; vertical D-MOS FET technology; the Siemens, power MOS also known as SIPMOS technology; DIMOS, i.e. "doppelt implantiert" MOS technology; HEX FET technology; and MISFET "Metal-Insulator-Semiconductor"-technology with a multiple layer gate, containing an insulated layer.

Forming the quotient of tne maximum source-drain voltage ($V_{DS\,max}$) and the source-drain voltage of the transistor (V_{on}) in switched state ($V_{on} = I_{D\,max} R_{on}$, being

Table 6.2. Comparison of the Q Factors of MOS FET's with Those of the Bipolar Transistors

Type	$V_{DS\,max}$ (V)	$I_{D\,max}$ (A)	R_{on} (Ohm)	$Q_T = V_{DS\,max}/V_{on}$
MOS FET's				
B 300	100	12	0.2	42
2N 6657	60	10	1.6	3.7
2N 6658	90	10	3	3
VN 64GA	60	12.5	0.3	16
IRF 131	60	12	0.18	28
IRF 150	100	28	0.045	70
IRF 151	60	28	0.045	48
IRF 350	400	11	0.3	120
IRF 351	350	11	0.3	105
BUZ 10	50	12	0.1	42
BUZ 20	100	8	0.2	63
BUZ 34	200	14	0.2	71
BUZ 45	500	8.6	0.6	96
BUZ 54	1000	4.7	2.0	105
Bipolar BUX 80	800 (V_{CSM})	15 ($I_{C\,max}$)	1.5 V ($V_{CE\,sat}$)	530

Table 6.3. Major Data of HEX FET'S International Rectifier

Type	$V_{DS\,max}$ (V)	$I_{D,out}$ (A)	P_D (W)	$R_{DSon,\,max}$ V_{GS} = at 10 V (ohms)	Package
IRF 533	60	8	75	0.25	TO-220
IRF 151	60	28	150	0.055	TO-3
IRF 532	100	8	75	0.25	TO-220
IRF 150	100	28	150	0.055	TO-3
IRF 633	150	5	75	0.6	TO-220
IRF 231	150	7	75	0.4	TO-3
IRF 632	200	5	75	0.6	TO-220
IRF 230	200	7	75	0.4	TO-3
IRF 733	350	3	75	1.5	TO-220
IRF 351	350	11	150	0.3	TO-3
IRF 732	400	3	75	1.5	TO-220
IRF 350	400	11	150	0.3	TO-3
IRF 833	450	2.5	75	2	TO-220
IRF 431	450	3.5	75	1.5	TO-3
IRF 832	500	2.5	75	2	TO-220
IRF 430	500	3.5	75	1.5	TO-3

equivalent to the saturation voltage of a bipolar transistor), this quantity may be regarded as the Q_T factor of the switching transistor as it can be seen in Table 6.2. which summarizes these figures for a number of MOS FET's and a bipolar transistor. It may be seen that the MOS FET's have not yet attained the Q_T factor of bipolar switching transistors.

Table 6.3. gives a review of HEX FET's manufactured by IR (International Rectifier). The table lists only the transistors with minimum and maximum drain current for each voltage group.

6.3.3 Driving of the Switching Power Transistors

The high efficiency of switched-mode power supplies arises because there is a very low dissipation in the "on" as well as in the "off" states of the transistor, i.e. the product $V_{CE}I_c$ is relatively small. The quality of the transistor as a switch is determined by the saturation resistance, the leakage current and the change-over times. The actual power loss arises during the short change-over time: this is determined by the switching characteristics of the transistor and the pulse characteristic (slope) of the base current.

It is generally true that an optimum base drive reduces the load of the transistor, at the same time increases its loadibility.

Figure 6.12 shows the collector–emitter voltage and collector–current waveforms of the power transistor in a converter. An important point for switched-mode