SIEMENS

Miniaturization of Switched Mode Power Supplies by the Use of SIPMOS Small-Signal Transistors with High Blocking Capability

Until now, more extensive miniaturization of switched mode power supplies has been held up by the lack of suitable components. One definitive component for switched mode power supply systems is the power semiconductor switch, capable of withstanding direct mains voltages and with a reverse voltage of up to 800 V. The new SIPMOS small-signal transistors now meet these requirements.

The major systems integration challenge for the new generation of potential-free power supply devices is to increase the power density, with the objective of miniaturizing switched mode power supplies. On top of this, additional absolute requirements for future switched-mode power supplies are to increase system efficiency across the entire load range, to improve the electrical parameters, such as the accuracy of output voltage regulation, the dynamic regulation of interference factors etc., plus the reliability which largely determines the failure rate of modern electronic products in the communications industry. Furthermore, the current industry standard takes for granted a high level of operational security, such as protection against overload and short-circuiting, overvoltage and undervoltage protection, overtemperature protection, raised ambient temperature etc. In the past, attempts to build significantly smaller switched-mode power supplies have repeatedly failed because of the components required. One essential component for them is the power semiconductor switch, particularly for AC/DC converters which have a rectified mains voltage of 220 V on the input side and must be laid out for a reverse voltage of 800 V. The newly developed SIPMOS small-signal transistors, which in the SOT223 package have a reverse voltage of V_{Br} = 800 V, provide a suitable switch for this application.

Technology of the High-Voltage Small-Signal Transistor

The type BSP 300 small-signal transistor is a standard N-channel SIPMOS transistor which has a reverse voltage of 800 V, a switch-on resistance R_{DSon} = 20 Ω and a drain current of I_{D} = 0.19 A ($I_{\mathrm{D\text{-}Puls}}$ = 0.76 A). The transistor is produced using SIP-III technology, and at the same time its robustness has been improved. Accordingly, the two critical directions of further development in the MOS transistors continue to be the reduction in the switch-on resistance, R_{DSon} and further improvements in robustness. In the case of 800 V blocking MOS transistors, the switch-on resistance is determined – as **Figure 1a** shows – primarily by the doping and the thickness of the epitaxial region, and less by the cell geometry and the substrate region, so that it is only possible to achieve limited reductions in R_{DSon} by optimization of these technological parameters. By contrast, with 50 V blocking MOS transistors the switch-on resistance is determined primarily by the cell geometry and the thickness of the substrate.

The second main direction of development for MOS transistors is improvement of their robustness. This has been achieve both on the input side, by improved gate oxide with test voltage values of greater than 50 V (the datasheet only permits 20 V), and on the output side by a homogeneous cell design, so that these transistors are now avalanche-proof. The robustness of MOSFETs is often a decisive selection criterion for users. The additional circuitry depends on the energy which arises during switch-off. Whereas high-energy switch-off overvoltages can mostly be kept in bounds by suitable circuit layout, protecting against externally injected overvoltages is not so simple. Switch-off overvoltages arise not only in circuits with inductive loads, they are also triggered by the extremely short switching times of the MOSFETs themselves. The parasitic inductances, which are unavoidable in every circuit, can cause overvoltages to be generated during switch-off which exceed the breakdown voltage, and drive the transistor into breakdown. Independently of this, there is a danger of the injection of voltage transients from outside the circuit. In both cases, the energy which is released in the MOSFET can – depending on the amount of energy – lead to failure of the component.

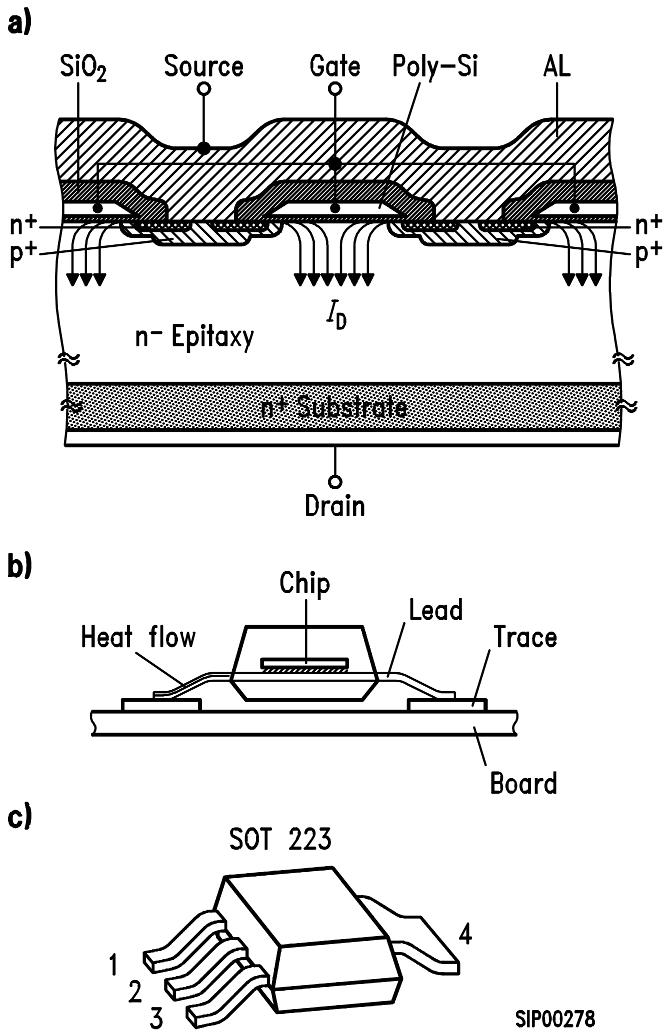

Figure 1
Technology and Structure of the 800 V Blocking Small-Signal Transistor
a) Semiconductor Structure of the Small-signal Transistor.
b) Structure of the Transistor Module in the Package: Heat is Conducted to the Board via “Lead” contacts and “Traces”
c) Structure of an SOT223 Package with Contact Pins

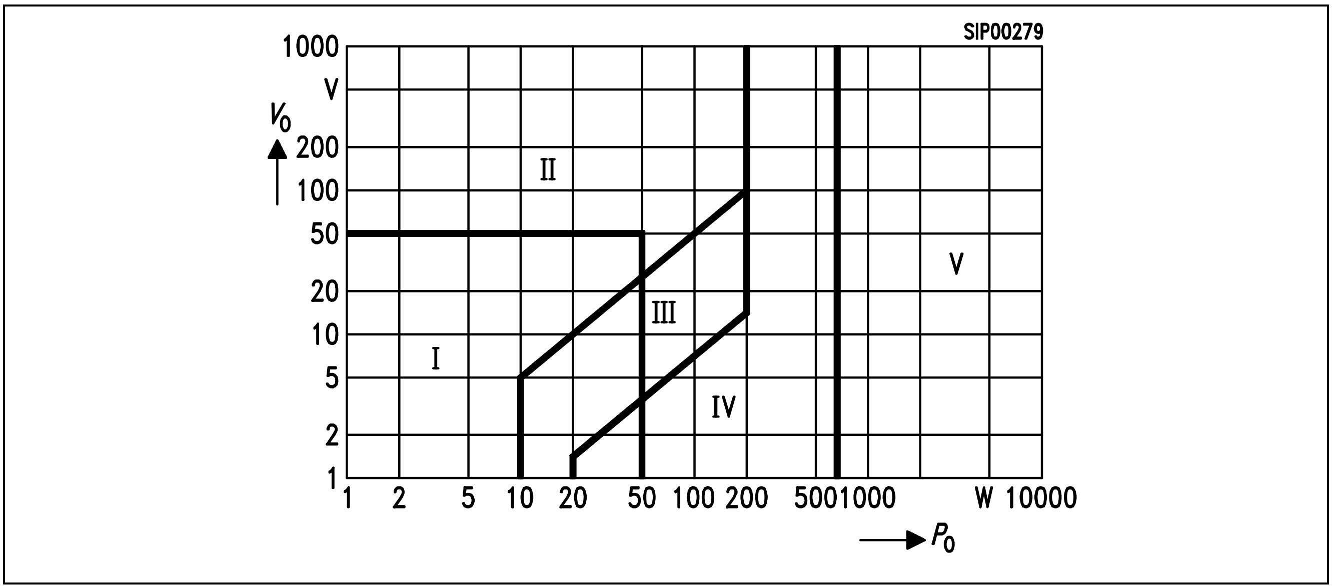

Figure 2
Input Voltage and Power Range Requirements Covered by the Various Switched-mode Power Supply Technologies. This Article is Concerned with Region II

SMD Package – An Important Step Towards System Optimization

With the SOT223 package, the system is firstly more robust and reliable, and secondly it is more compact and performs better. However, cost-effective solutions with high power densities call for automated insertion of SM components into PCBs.

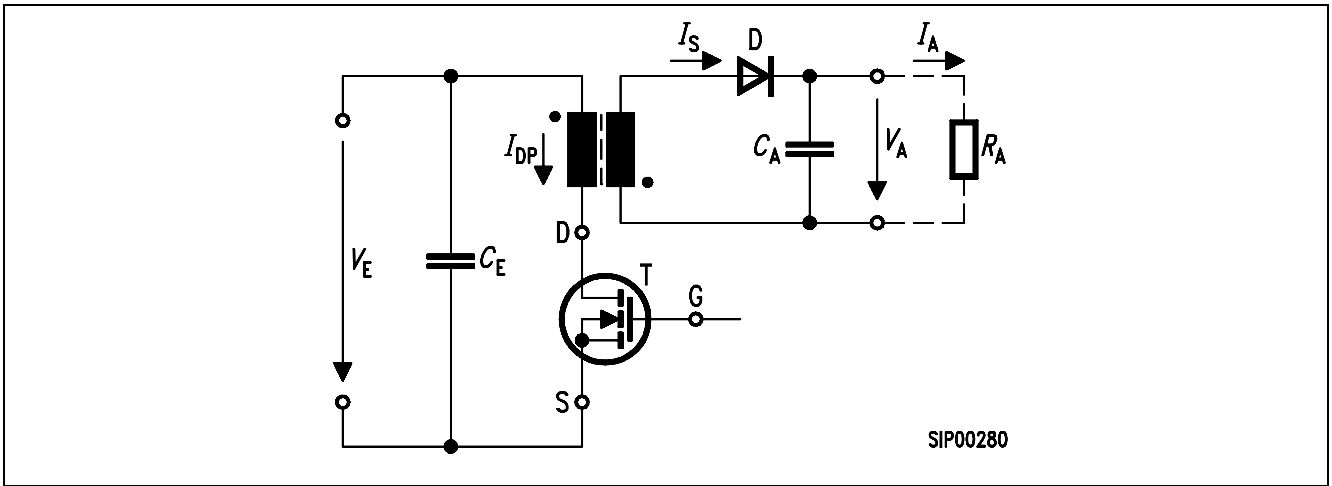

Figure 3
Circuit Diagram of the Flyback Converter: the Power Transistor Switches the Current I_{DP} on the Primary Side

Conventional power packages, such as the TO220 variants with large cooling fins, bring only little advantage unless mounted directly on a heat sink. With the SOT223 package – as shown in **Figure 1b** and **Figure 1c** – some of the heat flows through the molding material and is given up to the environment by radiation and convection, but most of the heat flows away through the connecting pins. Depending on the nature of the p.c.b. mounting, this can achieve heat conduction values of $R_{\text{th-JA}}$ = 60 to 70 K/W. These values relate to an epoxy p.c.b. of 40 mm $\times$ 40 mm $\times$ 1.5 mm, with 6 cm^2 area of Cu.

This allows power loss up to:

$$P_{\text{tot}} = \frac{T_{\text{j}} - T_{\text{A}}}{R_{\text{thJS}} - R_{\text{thSA}}}$$ **equation 1**

J = Junction
S = Soldering point
A = Ambient

$$P_{\text{tot}} = \frac{150\ ^\circ\text{C} - 25\ ^\circ\text{C}}{9\ \text{K/W} + 63\ \text{K/W}} = 1.7\ \text{W}$$ **equation 2**

Using the BSP 300 in a Switched-Mode Power Supply

A typical application for a high-blocking small-signal transistor – and the BSP 300 has been optimized as such – is its use in a switched-mode power supply. All electronic devices require a supply of one or more DC voltages for low loads. Supplying these devices from the AC mains network requires power packs which produce these DC voltages from the AC mains and, at the same time, ensure galvanic isolation from the input. A switched-mode power supply works on the principle that the mains input voltage is first rectified, filtered and this DC voltage is then pulsed and stepped down by a transformer. The transformer converts the pulsed DC voltage, which on the secondary side is then rectified again and filtered. A closed-loop control circuit keeps the output voltage constant by exercising control over the pulse duty ratio and/or the switching frequency of the power MOSFET.

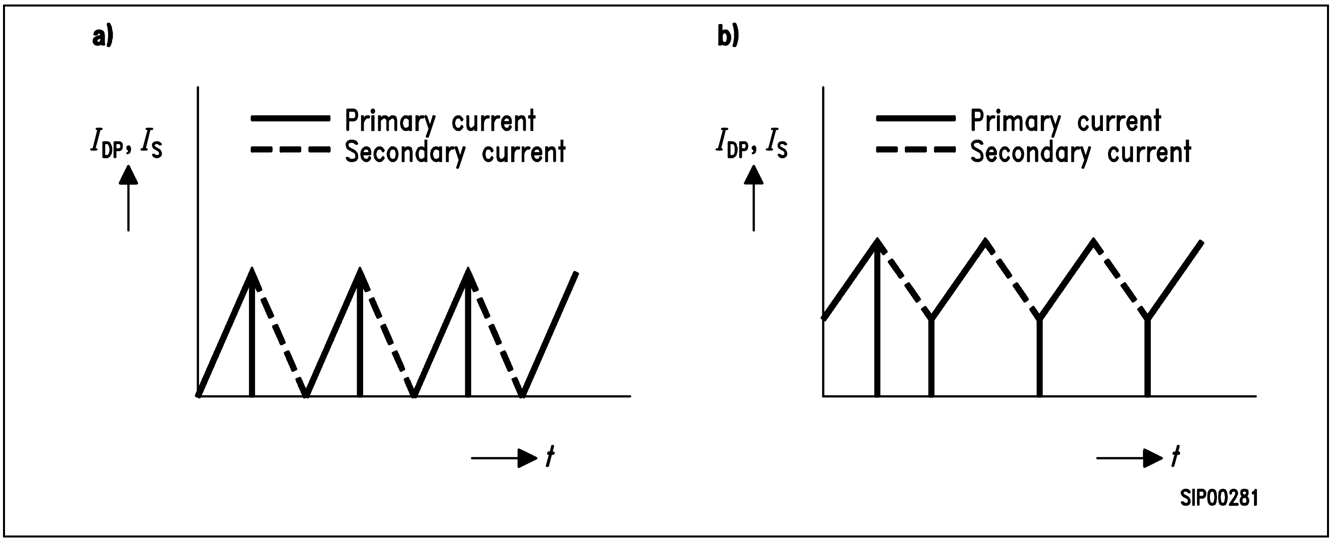

Figure 4
Schematic of Current Waveform in the Flyback Converter's Transformer
a) Delta Mode: at the Start of each new Cycle the Transformer Core is Demagnetized
b) Trapezoidal Mode: each new Cycle begins at a Point when there is still Energy in the Transformer Core

The great advantage of this compared to conventional power packs, apart from the low weight and volume, it the higher efficiency and lower heat dissipation because the losses are lower. A particular topology is chosen on the basis of the requirements to be met by the switched-mode power supply. The requirements covered by the various switched-mode power supply technologies is shown in **Figure 2**:

In region I (V_0 = 50 V, P_0 = 50 W) use is made mainly of constant-current transformers, provided that potential isolation is not required. These types of converter are least demanding in terms of switching technology, but have the disadvantage that a very large current flows through the switching transistor.

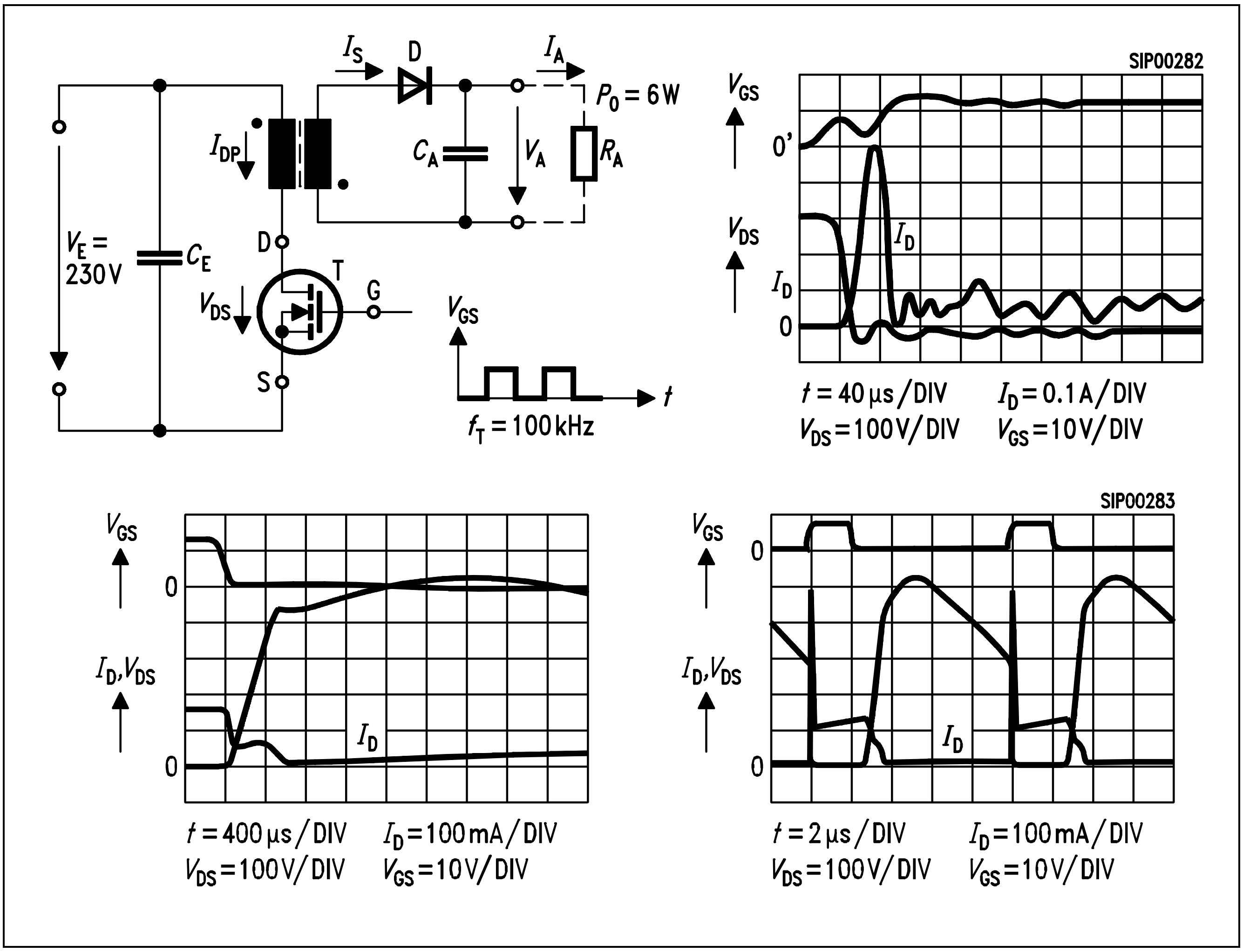

Figure 5
Experimental Results:
Circuit Diagram of the Flyback Converter and Waveforms of $v_{GS}(t)$, $i_D(t)$ and $v_{DS}(t)$ for f_T = 100 kHz; V_E = 230 V

Within region II, use is made mainly of flyback converters (for circuit topology see **Figure 3**), in region III both flyback converters and single-ended forward converters are used, in region IV single-ended forward converters, and in region V push-pull converters. This article is concerned exclusively with region II, in which flyback converters are customary for low power uses combined with high input voltages. This region encompasses a large proportion of the plug-in power supplies, such as chargers for mobile phones, hand-held computers, power packs for telecommunication devices and laptops, etc. In the case of the flyback converter topology (**Figure 3**) the transformer takes up energy during the transistor's conducting phase and, during the latter's blocked phase, transfers it to the output. By providing the transformer with appropriate poles, it is possible to generate output voltages of any required level and polarity. Flyback converters require only one magnetic component, and are less expensive than forward converters. They are use mainly for lower power and output voltages. Another advantage of the flyback converter topology is its galvanic isolation of the input and output, and the relative ease with which several galvanically isolated output voltages can be generated.

Table 1

Type	V_{DS}	I_D	R_{DSon}	Package	
BSP 300	800 V	0.19 A	20 Ω	SOT223	Available now
BSP	800 V	0.25 A	12 Ω	SOT223	Samples: early 97

For flyback converters, energy is transferred from the rectified mains voltage into the transformer during the switching transistor's conductive phase, and stored. No current flows on the secondary side, because diode D is oriented in the blocking direction. When the transistor is turned off, the voltages in the inductances reverse, because they attempt to maintain the current flow. The diode is now oriented in the direction of current flow, and energy is transferred to the output. Depending on the nature of the current waveform in the transformer, two operating modes are distinguished:

a) Delta Mode

Here, the current through the primary inductance has a triangular or delta-shaped waveform (**Figure 4a**). The triangular shape arises because at the start of each new cycle the core of the transformer is demagnetized. A distinction is made between continuous or gapped delta mode, depending on whether each new cycle starts immediately after demagnetization or there is a time gap before the start of the new cycle.

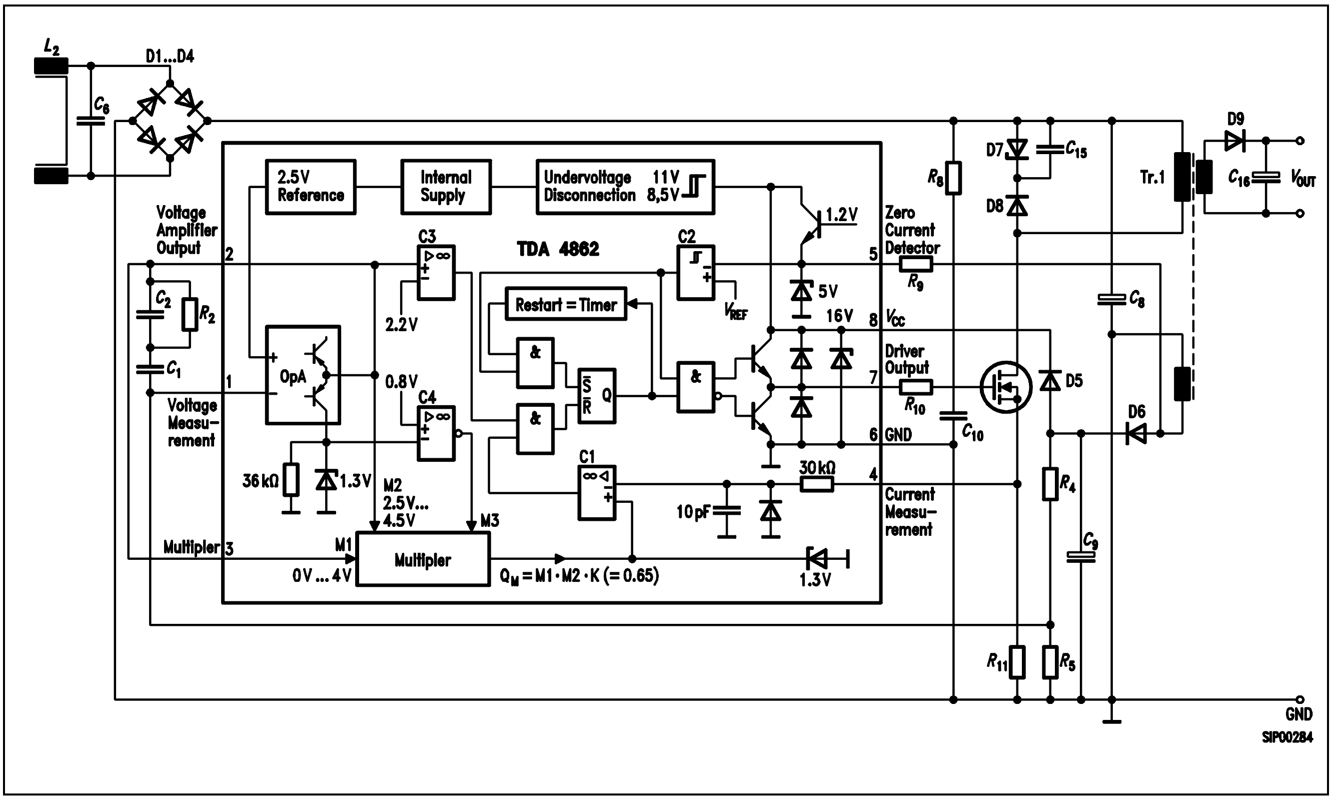

Figure 6
Block Diagram of the TDA 4862 Driver Module for Flyback Converters

b) Trapezoidal Mode

Here, the current through the primary inductance exhibits a trapezoidal waveform (**Figure 4b**). This arises because each new cycle begins even though there is still energy stored in the transformer's core.

The device used for the power switch in the flyback converter shown in **Figure 5** was a BSP 300. The flyback converter was pulsed at f_T = 100 kHz. The current and voltage waveforms, $v_{GS}(t)$, $i_D(t)$ and $v_{DS}(t)$, are also shown in **Figure 5**. The switched mode power supply was then operating in trapezoidal mode.

The following relationship is relevant for determination of the power loss:

$$\frac{T_j - T_A}{R_{thJS} - R_{thSA}} = P_{tot} \leq R_{DSon} \times I_D^{2} \times D + V_i \times \frac{t_r + t_f}{2} \times I_d \times f_T \qquad \textbf{equation 3}$$

Conducting losses   Switching losses

t_r = Current rise time at switch-on

t_f = Current decay time at switch-off

D = Pulse duty ratio

R_{thJS} = 9 K/W

R_{thSA} = 63 K/W

$\Rightarrow P_{tot}$ = 1.7 W dissipatable loss for the specified configuration

i_D = 150 mA

For an input voltage of V_{CE} = 230 V and a drain current of i_D = 150 mA, this gives:

P_{Trges} = 0.26 W (Conducting losses) + 0.6 W (Switching losses) = 0.86 W = P_{tot} **equation 4**

For an input voltage of V_{CE} = 230 V and a drain current of i_D = 250 mA, this gives:

P_{Trges} = 0.65 W (Conducting losses) + 1.05 W (Switching losses) = 1.7 W = P_{tot} **equation 5**

This indicates that the BSP 300 in an SOT223 package can be used, in the configuration shown with V_{CE} = 230 V and a current of up to 250 mA, in a flyback converter operating in trapezoidal mode at up to f_T = 100 kHz.

Application areas for this transistor are chargers/power packs for 220 V mains supplies, working in fixed frequency flyback converters pulsed at f_T = 100 kHz (with reverse voltage impulses up to 700 V) and delivering an output power of P_0 = 6 W (palmtops) and P_0 = 12 W (laptops). To reduce the loss power and increase the power range, another 800 V blocking small-signal MOSFET is in development, with R_{DSon} = 12 Ω. Two high-blocking small-signal MOSFETs will then be available (see **Table 1**).

IC Driver for the Flyback Converter

To drive the flyback converter shown in **Figure 5**, a special integrated driver circuit has been developed (**Figure 6**). This IC driver (**TDA 4862**) drives the flyback converter in such a way that it draws a sinusoidal current from the mains supply and provides a regulated DC supply at the output. The transistor does not switch on until the current in the diode falls to 0. This results in a high-frequency delta-shaped current flow in the choke. The undervoltage monitoring mechanism, which typically has a switch-on threshold of 10.5 V and a switch-off threshold of 8.5 V, ensures that the switching circuit is operating completely reliably before the driver output is released. The driver output is designed for driving power MOSFETs with a current carrying capacity of up to ± 500 mA. To avoid feedback, the driver output has clamping diodes to ground and the supply voltage, with a current carrying capacity of 100 mA. The control amplifier compares the stepped-down output voltage at its inverting input against a reference voltage of 2.5 V which is highly-accurate across the full temperature range. Via its inverting input, the current comparator detects the voltage drop across the voltage resistor located in the power MOSFET’s sourcepad. The voltage spikes which arise during rapid switching of the transistor are suppressed by the integral low-pass filter. The detector determines the point in time at which the current in the diode has dropped to zero, and then triggers a new cycle. After the current comparator has initiated the switch-off process, and the power MOSFET is storing charge, the diode takes over current conduction.