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Application Note

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TDA4862

TDA4862 - Technical Description

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Power Management & Supply

Infineon technologies

Never stop thinking

Contents:

Short Description

The TDA 4862 integrated circuit controls a boost converter in a way that sinusoidal current is taken from the single-phase line supply and stabilized DC voltage is available at the output. The circuit acts as a harmonic filter which limits the harmonic currents resulting from the pulse charge currents of the capacitor during rectification in a conventional capacitive input rectifier circuit. The power factor which describes the ratio between active and apparent power is almost 1 and line voltage fluctuations are compensated very efficiently, as well.

Technical Description TDA4862

Control Method

The control method of the harmonic filter is based on the physical relationship between current and voltage at the boost converter choke. The transistor does not switch on until the current in the boost converter diode turns zero. This creates triangular currents at a high frequency in the choke as it is principally shown in figure 1, avoiding high-loss reverse recovery currents of the diode. If triangular currents flow through the boost converter choke uninterruptedly, the mean input current calculated over a high-frequency period is exactly half as high as the peak value of the high frequency choke current. If the peak values of the choke current are on an envelope which is proportional to a sinusoidal low-frequency input voltage, a sinusoidal input current will be drawn from the mains after smoothing by means of an RFI suppression filter. The RFI suppression filter is designed in a way that the valid EMI limits at the inputs are not exceeded. Using this control method, the operating frequency of the active harmonic filter changes with the input voltage and the load.

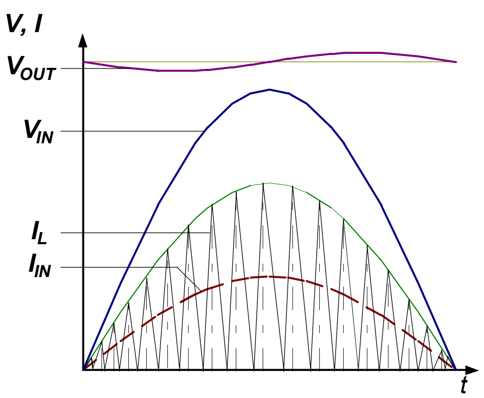

Figure 1: Electrical input parameters and choke current at discontinuous conduction mode operation

Characteristics

Power Supply and Self-Start

An undervoltage lockout with a turn-on threshold of 11 V typically and a turn-off threshold of 8.5 V typically assures that the IC is functional before the driver output is enabled. In the stand-by state prior to enabling the driver the IC consumes a current of less than 0.2 mA. A startup timer generates a set of pulses for the turn-off flip-flop, if the driver output stays in low state levels for longer than 150 µs. In order to guarantee safe supply from a current source the supply voltage pin 8 is internally limited to 17 V to ground. Thus, the IC has all functions necessary for low-loss self-start.

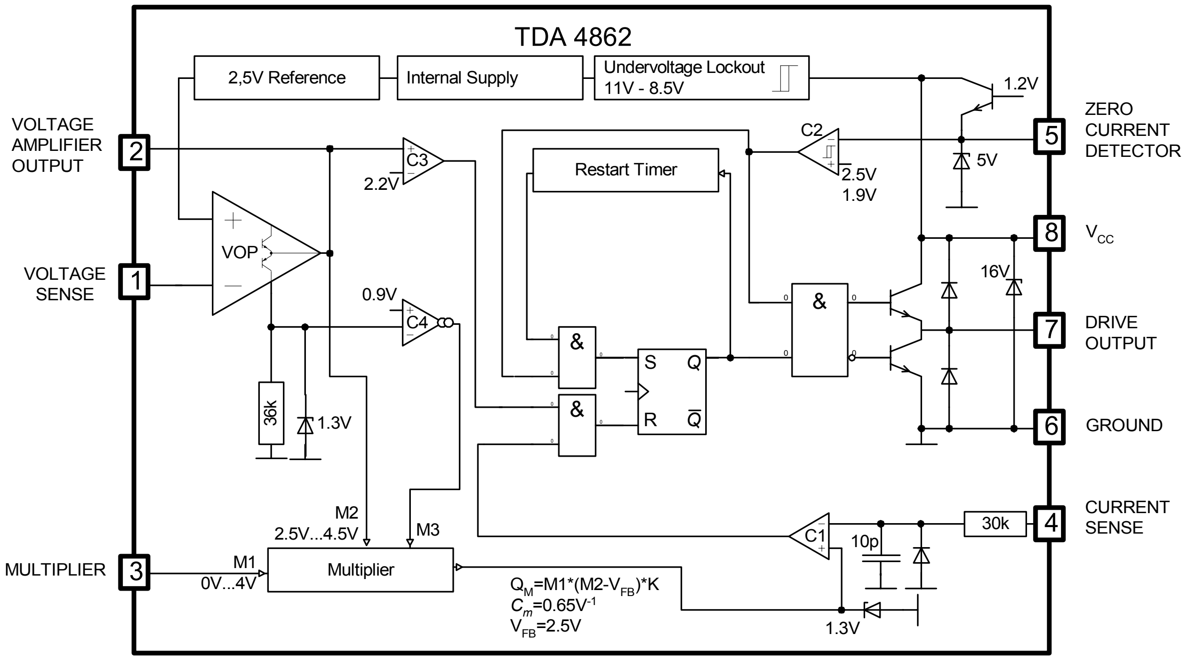

Figure 2: Scheme of TDA4862

Driver output

The driver output has been designed to drive power MOSFET with a current capability of ± 500 mA. In order to avoid reverse currents the driver output is equipped with clamping diodes connected to ground and supply voltage with a current rating of 100 mA. In standby state the driver output actively asserts a LOW level with a residual voltage of 1.5 V and 5 mA dissipation current.

Control amplifier

The control amplifier compares the divided output voltage at its inverting input with a highly accurate reference voltage of 2.5 V, with a maximum deviation of less than ±2% over the total temperature range (–40°C < T_J < 150°C), at its non-inverting input. For the purpose of control loop compensation a feedback network is inserted between the amplifier output (pin 2) and its inverting input (pin 1). A feedback design using only one capacitor as an I-controller causes oscillating transient response, because the boost converter, as a controlled current source, with the storage capacitor at its output delays the phase by almost 90° in no-load and in low-load operation. The transient response is more favorable if the control amplifier is designed as a PIT1-controller (see design steps).

The output voltage of the control amplifier ranges from 0.9 V to 4.3 V and can be loaded with a current of 1 mA (source) and 2 mA (sink), respectively. The output voltage of the control amplifier is monitored by a comparator. If the output voltage drops 0.3 V below the reference level of 2.5 V (i.e. reference voltage) of the M2 multiplier input the driver output will be blocked directly via the turn-off flip-flop. This measure guarantees the stability of the output voltage in complete no-load operation, without interferences from offset voltages at the multiplier output or at the comparator input.

The output DC voltage of the boost converter is superimposed by double the mains frequency AC voltage ripple. The amplitude of the superimposed AC voltage depends on the capacity of the storage capacitor and the load. The superimposed AC, which is also fed back via the control amplifier, causes an undesirableed modulation of the line current drawn. Therefore the bandwidth of the controll amplifier is chosen which is considerably lower than twice the mains-frequency. However, this causes the controller to react slowly to sudden load changes which results in voltage overshoots and output breakdowns.

Overvoltage control

If at the boost converter output a higher voltage than the rated output voltage is generated as a result of voltage transients or load rejection, a current flows back from the output voltage divider to the operational amplifier output via the feedback network. This is shown in figure 3. The current *ΔI* is measured and in case of a threshold of 30 µA (typ.) is exceeded the multiplier output is controlled to zero potential via a third input M3. This measure causes the input current to be continuously

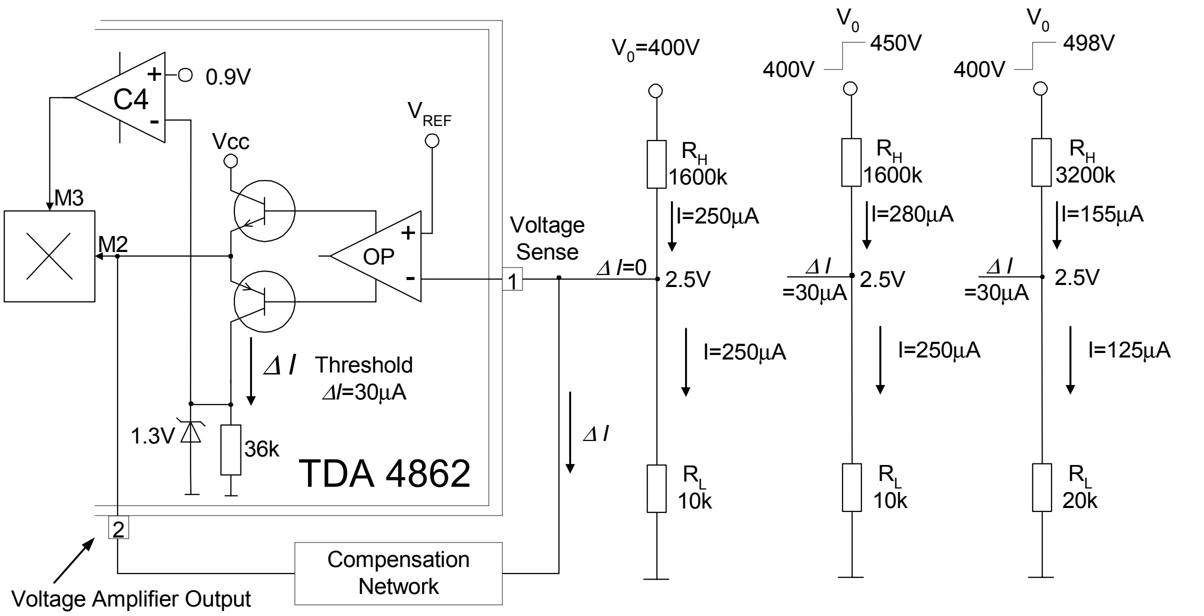

Figure 3: Examples of the output voltage divider

compensated back, thus avoiding uncontrolled oscillations of the line current drawn, as they usually appear with digital measures.

The switch-off level of the overvoltage control can be adjusted via the internal resistance of the output voltage divider. In normal operation state the voltage at the tap of the divider is 2.5 V (i.e. reference voltage). In case of higher than rated output voltage the excess divider current flows from the tap to the operational amplifier output via the feedback network. The overvoltage control is also guaranteed in the operational phases when the output voltage of the control amplifier reaches the upper limit threshold, because the dissipation current is measured as well. As soon as the output voltage of the control amplifier tends towards the minimum level, the comparator turns off at a level of 2.2 V to guarantee safe no-load operation.

Multiplier

The multiplier generates the turn-off threshold of the current comparator giving consideration to the waveshape of the feed voltage. In a typical application the rectified and divided supply voltage is applied to the input M1 (pin 3). The output voltage of the control amplifier is applied at the input M2 which – under constant load and ideal conditions – appears as DC voltage without superimposed AC shares. At the output of the multiplier a signal in the wave form of the rectified voltage corresponding to input M1 is generated which can be modified in its amplitude via the DC voltage at input M2. Superimposed AC voltage shares at the input M2 cause an undesired modulation of the line current drawn, unless they are part of the dynamic control processes. The level control range of the input M1 is 0 V to 4.0 V, the reference level being 0 V. The level control range of the input M2 is 2.5 V to 4.5 V, the reference level being 2.5 V. For multiplication a further, constant factor C_m = 0.65 V^{-1}, which is an internal factor of the multiplier, is effective. Its dimension is V^{-1} in order to comply with the following equation. In this way the current comparator level can be calculated as $V_{Qm} = C_m\ (V_{pin2} - V_{ref})\ V_{pin3}$.

The output voltage of the multiplier is limited to 1.3 V. This measure causes a defined turn-off threshold for current limitation. In this way, dangerous excess currents are avoided which can arise in particular in the case of an expanded input voltage range because the multiplier with its restricted dynamics re-stabilized the current consumption.

Current comparator

The current comparator detects the voltage decline at the shunt which is in the source path of the power MOSFET via its inverting input (pin 4) and which should have an intrinsic inductance as low as possible. When switching on the transistor voltage, spikes are generated at the shunt as a result of the intrinsic inductance of the shunt with turn on and the influence of the driver currents. An integrated low-pass filter suppresses these voltage spikes. As soon as the voltage decline at the shunt reaches the turn-off threshold defined by the multiplier, the turn-off flip-flop is reset and the driver switches off. The turn-off flip-flop prevents multiple pulses during the switching waveform of the power MOSFET. The turn-off delay time between comparator input and driver output is below 250 ns.

Detector

The detector finds the point of time when the current in the boost converter choke turned zero and then enables the control of a new pulse cycle. After the current comparator triggers the turn-off process and the power MOSFET blocks, the boost converter diode takes over the current. In this case the polarity of the voltage at the choke winding changes in a way that now a higher level voltage levell (V_{out}) is available at the drainside terminal of the choke compared with the mains rectifier side terminall (level V_{in}) of the choke. As soon as the choke current reaches zero and the boost converter diode blocks, the voltage reverses at the drain side terminal of the choke. The voltage at the choke winding turns zero or changes polarity. A second winding (detector winding) on the choke, which has approximately 1/5 of the number of turns compared with the mains winding, permits the change of polarity of the choke voltage to be registered without detrimental influences. Evaluation is effected by the detector function (pin 5) of the IC, with the drain side polarity of the detector winding being measured by means of a hysteresis-determined comparator.

The level for the acceptance of the „MOSFET blocks“ command from the turn-off flip-flop and for setting the flip-flop is 2.5 V (i.e. reference voltage) with rising voltage. In case of a voltage decline, which signals the zero crossing of the current, the switching level enabling the driver stage is 1.9 V. The voltage of the detector winding is applied to pin 5 via a high-ohmic resistance (10k to 47k). Clamping structures are available in the IC which limit the voltage at the input to +5 V and +0.6 V, respectively, at 10 mA maximum.

There are cases in which there is no significant detector signal to set the turn-off flip-flop. This may be the case when the supply voltage is switched on, in case of line overvoltage exceeding the output voltage and in no-load and low-load operation, when the voltage controlller specifies intermittant operation. In that case a startup generator is activated which supplies a set of pulses to the turn off flip-flop if the driver output stays on LOW-level longer than 150 µs.

Applications of the TDA4862

The following applications demontrate the good performance of the TDA4862 controlling a power factor preconverter. The design steps indicate the method of the calculation of the components values. Lamp ballast designs as well as a design for switched mode power supplies (SMPS) are given here as examples. Circuit diagrams and measurement results at different operating conditions establish a good basis for evaluation.

The tables of page 17 ff. also consists of a column called I_Z which contains the values for the surplus current of the auxiliary power supply for the IC bypassed with a 15 V zener diode. The zener current indicates a suffucient IC supply. Therefore it should be low enough to avoid unnecessary losses. There may be also states of operation when the zener current reaches zero. Then the actual supply voltage V_{CC} of the IC is figured.

Usually a single stage RFI-filter does not accomplish the RFI-standards. Therefore multiple stage RFI-filters are designed into these applications as an example how to suppress resonant oscillations of these filters.

Discontinuous conduction mode always results in a high switching efficiency, because it avoids reverse recovery losses of the boost converter diode. A high power factor, low harmonics, a wide input voltage range and a feedback controlled output voltage are the most omportant features of a power factor preconverter. The TDA4862 contains all control and monitoring functions to meet these demands.

Design steps

Input and output section

Application			2L-Ballast	1L-Ballast	3L-Ballast	SMPS
Nominal input voltage	V_{innom}		120V AC	230V AC	277 V AC	90 V – 270 V
Minimum input voltage	V_{inmin}	$= V_{innom} - 20\%$	96V AC	184V AC	221 V AC	90 V AC
Maximum input voltage	V_{inmax}	$= V_{innom} + 20\%$	144V AC	276V AC	332 V AC	270 V AC
Minimum peak input voltage	$V_{inPkmin}$	$=\sqrt{2}\ V_{innin}$	136V	260V	313 V	127 V
Maximum peak input voltage	$V_{inPkmax}$	$=\sqrt{2}\ V_{inmax}$	204V	390V	470 V	382 V
Estimated minimum efficiency	η	= 0.9				
Output power	P_{out}	$= \eta\ P_{in}$	75W	53W	110 W	150 W
Maximum peak input current	$I_{inPkmax}$	$= 2\ P_{out} / (V_{inpmin}\ \ \eta)$	1.225A	0.453A	0.781A	2.625 A
Maximum high frequency peak current	$I_{PkmaxHF}$	$= 2\ I_{inPkmax}$	2.45A	0.906A	1.562 A	5.25 A
Maximum current sense threshold	$V_{ISensemax}$	= 1.3 V				
Shunt resistor	R11	$= V_{ISensem} / I_{PkmaxHF}$	0.53Ω	1.44Ω	0.83 Ω	0.25 Ω
Nominal output voltage	V_{out}	Recommended minimum: $V_{inPkmax}$+30V	230 V DC	410V	480V DC	410 V DC
Reference voltage	V_{ref}	= 2.5 V				
Controller current at pin VAOUT	I_{VAOUT}	= 30 µA				
Output voltage divider	R5	$= V_{ref}\ /\ I_{R5}$	10k	10k	10k	10k
(Select I_{R5} = 250 µA)	R4	$= R_5\ (V_{out} - V_{ref}) / V_{ref}$	910k	1640k	1910k	1640k
Overvoltage threshold	V_{OV}	(Recommended: $1.1 V_{out}$)	257 V	462 V	537 V	462 V

Multiplier section

Application			2-Lamp-Ballast	1L-Ballast	3L-Ballast	SMPS
Multiplier inputs M1 and M2 dynamic voltage range		V_{m1R} = 3.8 V; V_{m2R} = 4.5V – V_{ref} = 2V				
Multiplier output limitation	V_{Qmmax}	=$V_{ISensemax}$ = 1.3V				
Multiplier gain	C_m	= 0.65				
V_{m1}(@ V_{Qm} = 1.3V; V_{m2R} = 2V)		=1.3V / (2V* C_m) =1V				
From multiplier output characteristic						
V_{m1lim}(@ V_{Qm} = 1.3V; V_{m2R} = 2V) = 1.2 V						
Select V_{m1} = V_{m1lim} = 1.2 V		@ $V_{inPkmin}$ =	136 V	260V	313V	127V
Select upper resistor of input voltage divider	R6		1M	2M	2M	940K
Lower resistor of input voltage divider	R7	=R6·V_{m1lim} / ($V_{inPkmin}$ – V_{m1lim})	8.89k	9.27k	7.69k	8.95k
Low pass filter capacitor	C4	=1 / (2π·R2·*f*) {1 kHz<*f*<3kHz}	10 nF	10 nF	10 nF	10 nF
Test: Input range: V_{m1}(@$V_{inPkmax}$) < V_{m1R} = 3.8 V ? otherwise select V_{m1}(@$V_{inPkmin}$) < V_{m1lim} = 1.2 V			1.80V=OK	1.80V=OK	1.80V=OK	3.62V=OK

Boost inductor section

In this section two different approaches for the calculation of the transformer primary inductance L_P are presented. The first one is recommended for a small input voltage range application or for applications with nearly constant output power, e.g. lamp ballasts. Therefore only one example is executed here. The other one is suitable for the demands of wide range applications like they occur in SMPS. All the values of the sections before are still valid.

<table>
<tr><th>2-Lamp-Ballast</th><th>SMPS-preconverter</th></tr>
<tr><td>
On-time of power switch: $T_{on} = L_P \cdot I_{PkmaxHF} / V_{in}$, $I_{PkmaxHF} = 2\ I_{inPkmax}$
Off-time of power switch: $T_{off} = L_P \cdot I_{PkmaxHF} / (V_{out} - V_{in})$
Pulse frequency: $f_p = \frac{1}{T_{on} + T_{off}} = \frac{V_{in} \cdot (V_{out} - V_{in})}{V_{out} \cdot L_P \cdot I_{Pk\max HF}}$

<u>Design criterion:</u>
Calculate L_P according to desired range of pulse frequency (e.g. 80 kHz < f_p <110 kHz) at nominal input voltage V_{innom} and rated output power P_{out}
$L_P = \frac{V_{innom} \cdot (V_{out} - V_{innom})}{V_{out} \cdot f_p \cdot I_{Pk\max HF}} = \frac{V_{innom} \cdot (V_{out} - V_{innom}) \cdot \eta \cdot V_{innom}}{V_{out} \cdot f_p \cdot 2P_{out}} =$
$= \frac{120\ V \cdot (230\ V - 120\ V) \cdot 0{,}9 \cdot 120\ V}{230\ V \cdot 90\ kHz \cdot 2 \cdot 75\ W} = 459\ \mu H$

Also possible:
Calculate L_P by selecting the on-time T_{on} in the range of 3 µs < T_{on} < 6 µs
$L_P = \frac{T_{on} \cdot V_{innom}}{I_{Pk\max HF}} = \frac{T_{on} \cdot V^2_{innom} \cdot \eta}{2 \cdot P_{out}} = \frac{5\ \mu s \cdot (120V^2) \cdot 0{,}9}{2 \cdot 75\ W} = 432\ \mu H$
Both inductances will be appropriate.
</td><td>
<u>Design criterion:</u>
Calculate L_P in order to obtain pulse frequencies higher than 25 kHz at maximum peak input voltage and twice of nominal output power and on minimum peak input voltage and twice of nominal output power
$L_P < \frac{V^2_{inPk\max} \cdot (V_{out} - V_{inPk\max}) \cdot \eta}{V_{out} \cdot f_p \cdot 2 \cdot 2P_{out}} = \frac{(382V)^2 \cdot (410\ V - 382\ V) \cdot 0{,}9}{410\ V \cdot 25\ kHz \cdot 2 \cdot 2 \cdot 150\ W} = 598\ \mu H$
and
$L_P < \frac{V^2_{inPk\min} \cdot (V_{out} - V_{inPk\min}) \cdot \eta}{V_{out} \cdot f_p \cdot 2 \cdot 2P_{out}} = \frac{(127V)^2 \cdot (410\ V - 127\ V) \cdot 0{,}9}{410\ V \cdot 25\ kHz \cdot 2 \cdot 2 \cdot 150\ W} = 668\ \mu H$

We therefore select L_P < 598 µH
</td></tr>
</table>

Application		Example
Ballast, V_{innom} = 120 V	$L = \frac{(120V)^2 \cdot (230V - 120V) \cdot 0.9}{230V \cdot 90kHz \cdot 2 \cdot P_{OUT}} = \frac{34{,}4mH \cdot W}{P_{OUT}}$	P_{OUT} =75 W L = 459 µH
Ballast, V_{innom} = 230 V	$L = \frac{(230V)^2 \cdot (410V - 230V) \cdot 0{,}9}{410V \cdot 90kHz \cdot 2 \cdot P_{OUT}} = \frac{116mH \cdot W}{P_{OUT}}$	P_{OUT} = 55 W L = 2,1 mH
Ballast, V_{innom} = 277 V	$L = \frac{(277V)^2 \cdot (480V - 277V) \cdot 0{,}9}{480V \cdot 90kHz \cdot 2 \cdot P_{OUT}} = \frac{162mH \cdot W}{P_{OUT}}$	P_{OUT} = 110W L = 1,47mH
SMPS, V_{in} = 90 V – 270 V	$L = \frac{90mH \cdot W}{P_{OUT}}$	P_{OUT} = 150W L = 600µH

Operating frequency f_p *versus peak input voltage* V_{inPk} *at constant output power* P_{out}

$$f_p(V_{inPk\max}) = \frac{V_{inPk\max} \cdot (V_{out} - V_{inPk\max})}{V_{out} \cdot L_P \cdot 2 \cdot I_{inPk\max}} = \frac{V_{inPk\max} \cdot (V_{out} - V_{inPk\max}) \cdot V_{inPk\max}}{V_{out} \cdot L_P \cdot 2 \cdot 2P_{in}} = \frac{V^2{}_{inPk\max} \cdot (V_{out} - V_{inPk\max}) \cdot \eta}{V_{out} \cdot L_P \cdot 4 \cdot P_{out}}$$

Operating frequency

$$f_p(\omega t) = \frac{V_{in} \cdot \sqrt{2} \cdot \sin\omega t \cdot (V_{out} - V_{in} \cdot \sqrt{2}\sin\omega t)}{V_{out} \cdot L_P \cdot 2 \cdot I_{in} \cdot \sqrt{2} \cdot \sin\omega t} = \frac{V_{in} \cdot (V_{out} - V_{in} \cdot \sqrt{2}\sin\omega t)}{V_{out} \cdot L_P \cdot 2 \cdot I_{in}} = \frac{(V_{in})^2 \cdot \eta}{V_{out} \cdot L_P \cdot 2 \cdot P_{out}}(V_{out} - V_{in} \cdot \sqrt{2} \cdot \sin\omega t)$$

Example

$$f_p(\omega t) = \frac{(120V)^2 \cdot 0.9}{230V \cdot 450\mu H \cdot 2 \cdot 75W} \cdot (230V - 120V \cdot \sqrt{2} \cdot \sin\omega t)$$

Figure 4 shows the pulse frequency dependent on the peak value of the input voltage for constant output power or constant primary inductance respectively. For example, the lower limit of the input voltage in wide range applications is about 0.3 V_{out}. The corresponding pulse frequency is then 40 % of the maximum pulse frequency. The upper limit in such applications is about 90 % of the output voltage V_{out}, which leads to a pulse frequncy of about 50 % of the maximal value.

It is important, that those two frequencies mentioned above are still above 25 kHz.

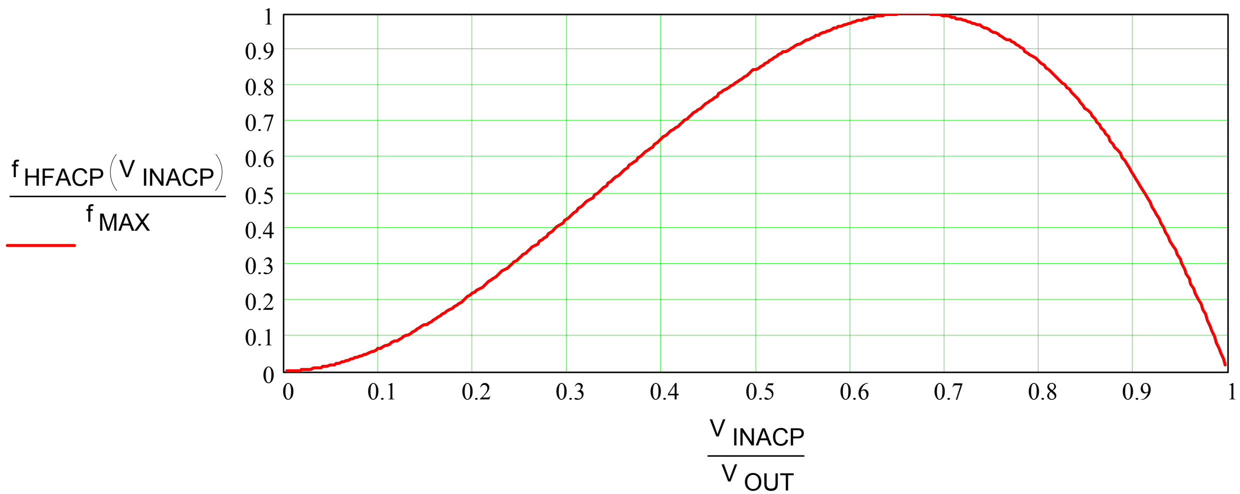

Figure 4: Pulse frequency f_p as a function of the input peak voltage

Output voltage controller:

Usually a PIT_1-design is used in PFC-circuits like it is shown in figure 8. The setting of the values of C1, C2 and R1 should hit the following targets:

- Good suppression of superimposed AC-share of the output voltage which has twice the frequency of the input voltage,
- good behaviour at load changes or changes of the input voltage,
- good behaviour at low load conditions.

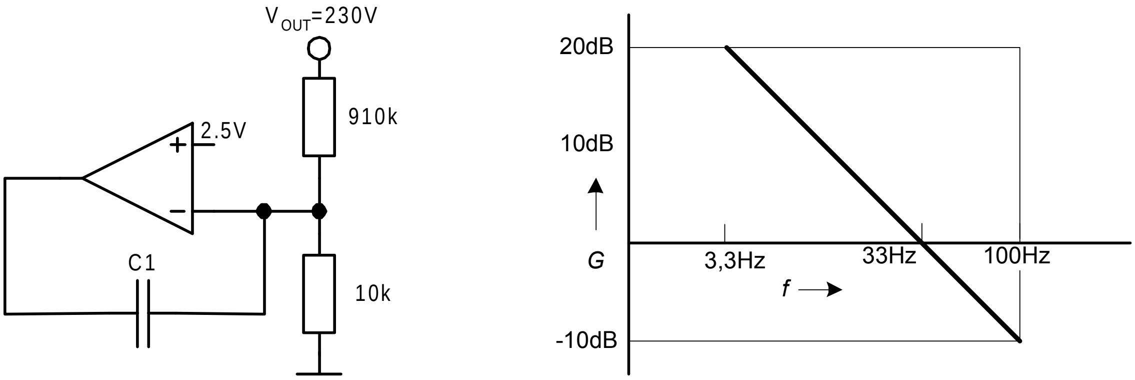

Figure 5: Output voltage controller with integral component

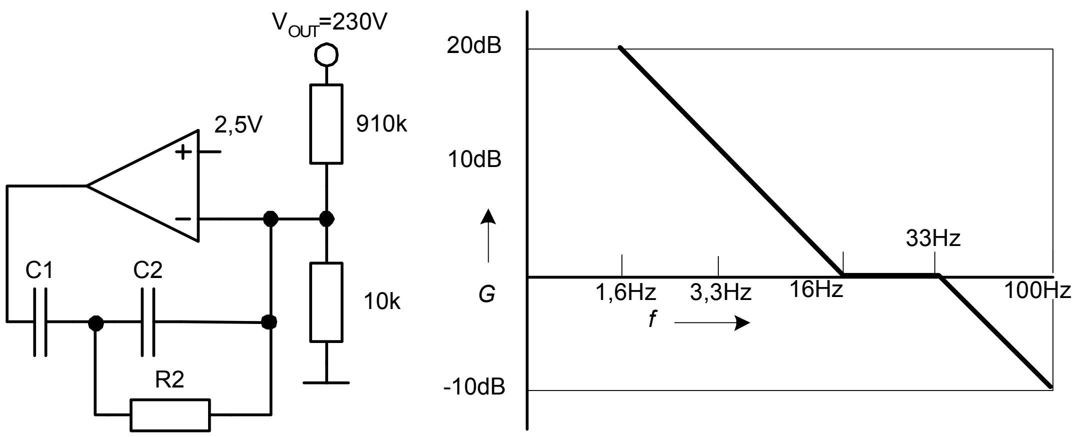

Figure 6: Output voltage controller in PIT1-design

Zero Current Detector

The upper threshold of the ZCD is max. 2.75V. For a continuous operation the difference between output voltage V_{out} and maximum input voltage $V_{inPkmax}$ and the transformation ratio of the inductor windings have to accomplish the following relation

$$(V_{out} - V_{inPk\max}) \cdot \frac{N_{ZCD}}{N_P} > 2.75V$$

The recommended transformation ratio of N_{ZCD}/N_P = 1/5 meets a minimal voltage difference of 14 V. If the detector input voltage doesn't achieve the upper threshold, the IC is operating with the timer frequency.

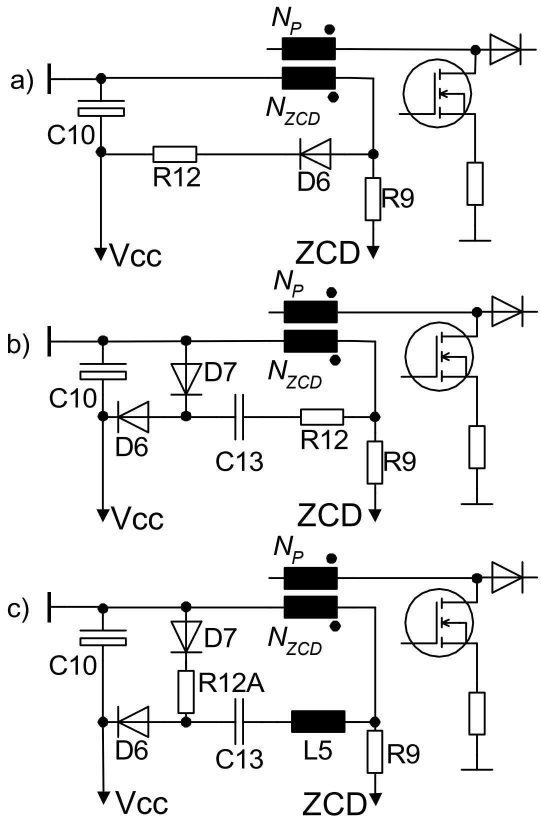

Figure 7: Auxiliary power supply realized with rectifier (a) and charge pump (b and c)

Auxilliary Power Supply

An obvious way to supply the IC is to use the detector winding. We have to care, that the supply circuit doesn't influence the detector signal. First, in a simple voltage mode supply, we use a diode, a storage capacitor C10 and a current limiting resistor R12. We achieve good results in ballast applications with the following design of the transformation ratio:

$$\frac{N_{ZCD}}{N_P} = \frac{V_{ZCD}}{V_{out} - V_{innom}}$$

V_{ZCD} = 22V...24V;
R_{12} = 220Ω...270Ω

Second in a charge pump supply, we use two diodes, two capacitors C10, C13 and one decoupling Resistor R12 or a decoupling inductor L5 (lower losses) and a current limiting resistor R12A, to avoid burn down at resonance frequency. This method of supply is to prefer in SMPS applications with wide input voltage range.

The supply current increases with the operating frequency at low load and is not dependend on the input voltage. We achieve good results with the following design of the transformation ratio:

$$\frac{N_{ZCD}}{N_P} = \frac{V_{ZCD}}{V_{out}}$$

$V_{ZCD} \approx 80V$,

C13 = 3nF...4nF
R12 = 390Ω...270Ω

Or C13 = 1 nF...1.5nF, L5 = 50 µH...100µH, R12A designed with C13 and L5 as a low-pass filter of Bessel characteristic.

Applications

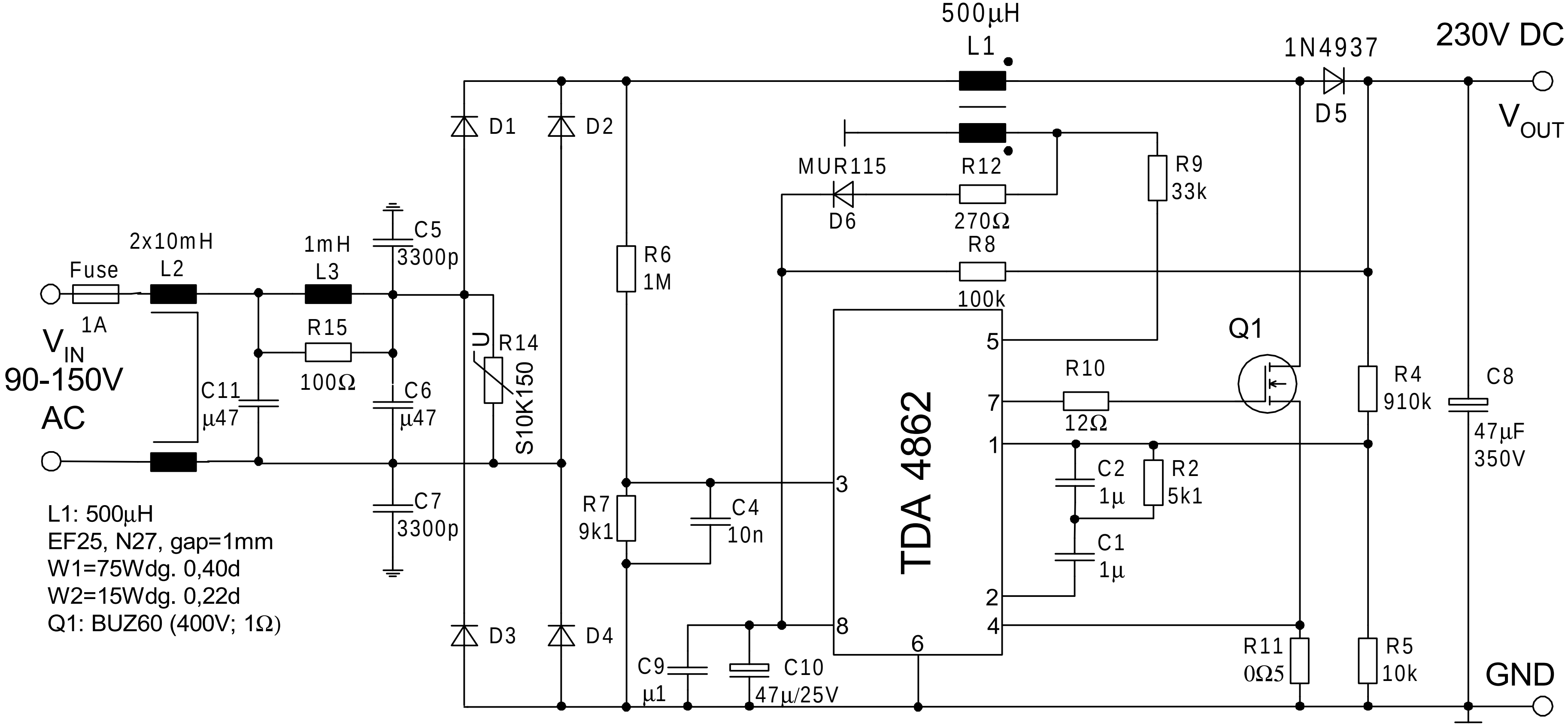

Figure 8: 75W Power Factor Preconverter with TDA 4862 and V_{innom} = 120V

Table 1: Measurement of input and output values

120V input for 2 x 35W lamp ballast (Cout = 47μF, L1=500μH)										
V_{RMS}	***I_{RMS}***	***P_{IN} real power***	***PF***	***THD***	***V_{OUT}***	***I_{OUT}***	***P_{OUT}***	***V_{OUTAC}***	***η***	***I_z(15V) or V_{CC}***
93V	0.882A	82.20W	0.999	2.0%	229V	0.328A	75W	25V	0.912	9.0mA
100V	0.812A	81.21W	0.999	2.5%	229V	0.328A	75W	25V	0.924	8.0mA
120V	0.663A	79.55W	0.999	3.2%	229V	0.328A	75W	25V	0.934	3.0mA
140V	0.563A	78.68W	0.997	4.3%	229V	0.328A	75W	25V	0.953	1.0mA
150V	0.524A	78.44W	0.996	4.7%	229V	0.328A	75W	25V	0.956	0.4mA
90V	0.392A	35.21W	0.998	3.0%	229V	0.124A	32.5W	13V	0.923	5.6mA
120V	0.289A	34.45W	0.993	5.0%	229V	0.124A	32.5W	13V	0.943	0.3mA
140V	0.249A	34.25W	0.984	6.5%	229V	0.124A	32.5W	13V	0.949	11.9V
90V	0.185A	16.54W	0.991	4.8%	229V	0.066A	15W	6V	0.907	1.7mA
120V	0.141A	16.32W	0.965	6.8%	229V	0.066A	15W	6V	0.919	11.7V
140V	0.124A	16.24W	0.933	9.5%	229V	0.066A	15W	6.5V	0.924	9.6V
120V	0.081A	8.65W	0.890	9.4%	229V	0.033A	7.5W	3V	0.867	9.8V
120V								30V		

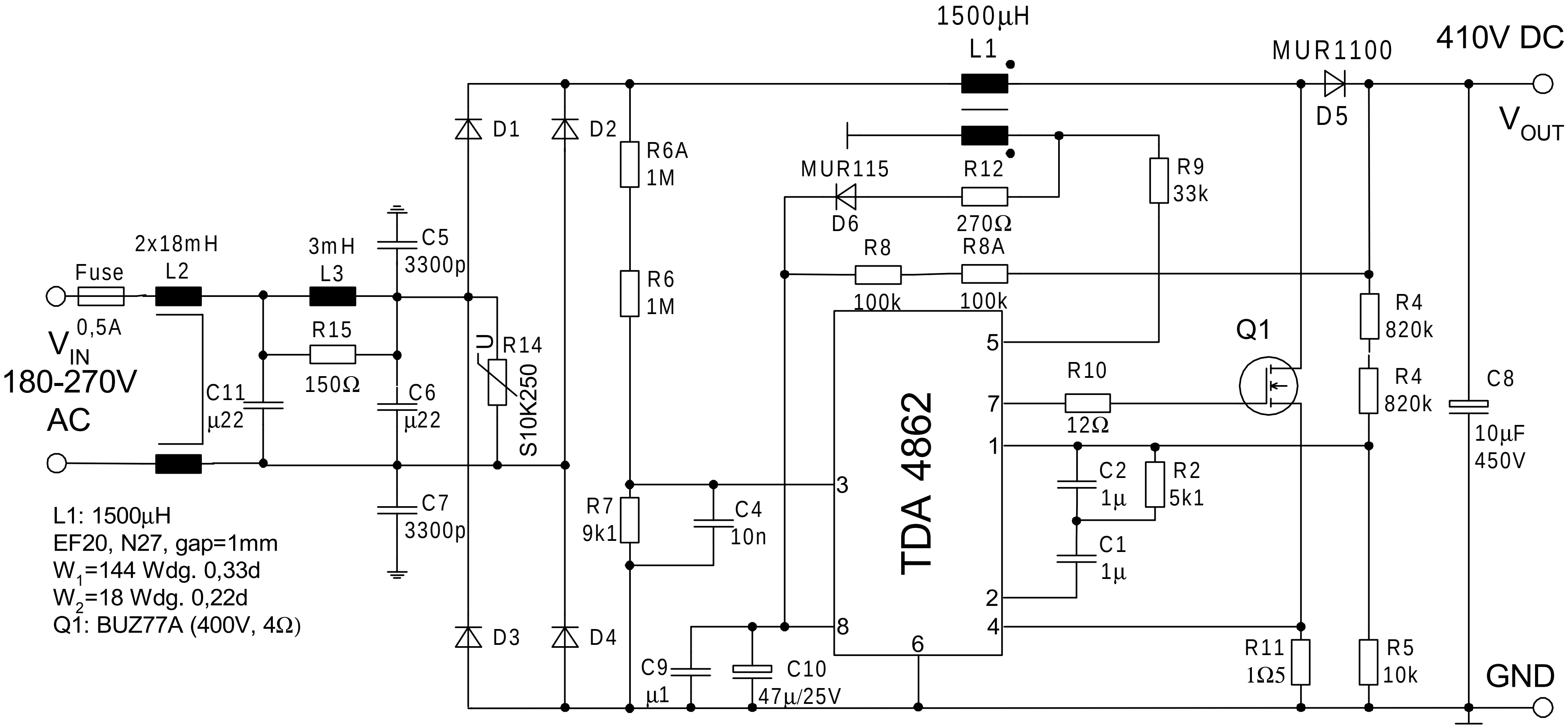

Figure 9: 53W Power Factor Preconverter with TDA 4862 and V_{innom} = 230V Input

Table 2: Measurement of input and output values

230V input for 50W lamp ballast (Cout = 10µF, L1=1.5mH)										
V_{RMS}	I_{RMS}	P_{IN} real power	PF	THD	V_{OUT}	I_{OUT}	P_{OUT}	V_{OUTAC}	η	I_z (15V) od. V_{CC}
180V	0.317A	57.16W	0.998	3.0%	409V	0.130A	53W	30V	0.927	9.9mA
200V	0.282A	56.38W	0.997	2.5%	409V	0.130A	53W	30V	0.940	6.9mA
230V	0.245A	56.02W	0.993	4.0%	409V	0.130A	53W	30V	0.946	4.0mA
250V	0.225A	55.76W	0.989	5.3%	409V	0.130A	53W	30V	0.950	2.8mA
270V	0.209A	55.61W	0.984	6.3%	409V	0.130A	53W	30V	0.953	1.9mA
180V	0.146A	29.36W	0.991	4.3%	409V	0.066A	27W	16V	0.920	7.6mA
230V	0.130A	28.95W	0.970	7.8%	409V	0.066A	27W	16V	0.933	1.8mA
270V	0.113A	28.8W	0.941	10.7%	409V	0.066A	27W	16V	0.937	0.2mA
180V	0.075A	12.68W	0.944	9.8%	409V	0.027A	11W	8V	0.868	4.0mA
230	0.063A	12.63W	0.865	12%	409V	0.027A	11W	8V	0.871	12.8V
270V	0.061A	12.60W	0.764	19%	409V	0.027A	11W	8V	0.871	12.0V
230V					409V		0	50V	0.871	10.8V

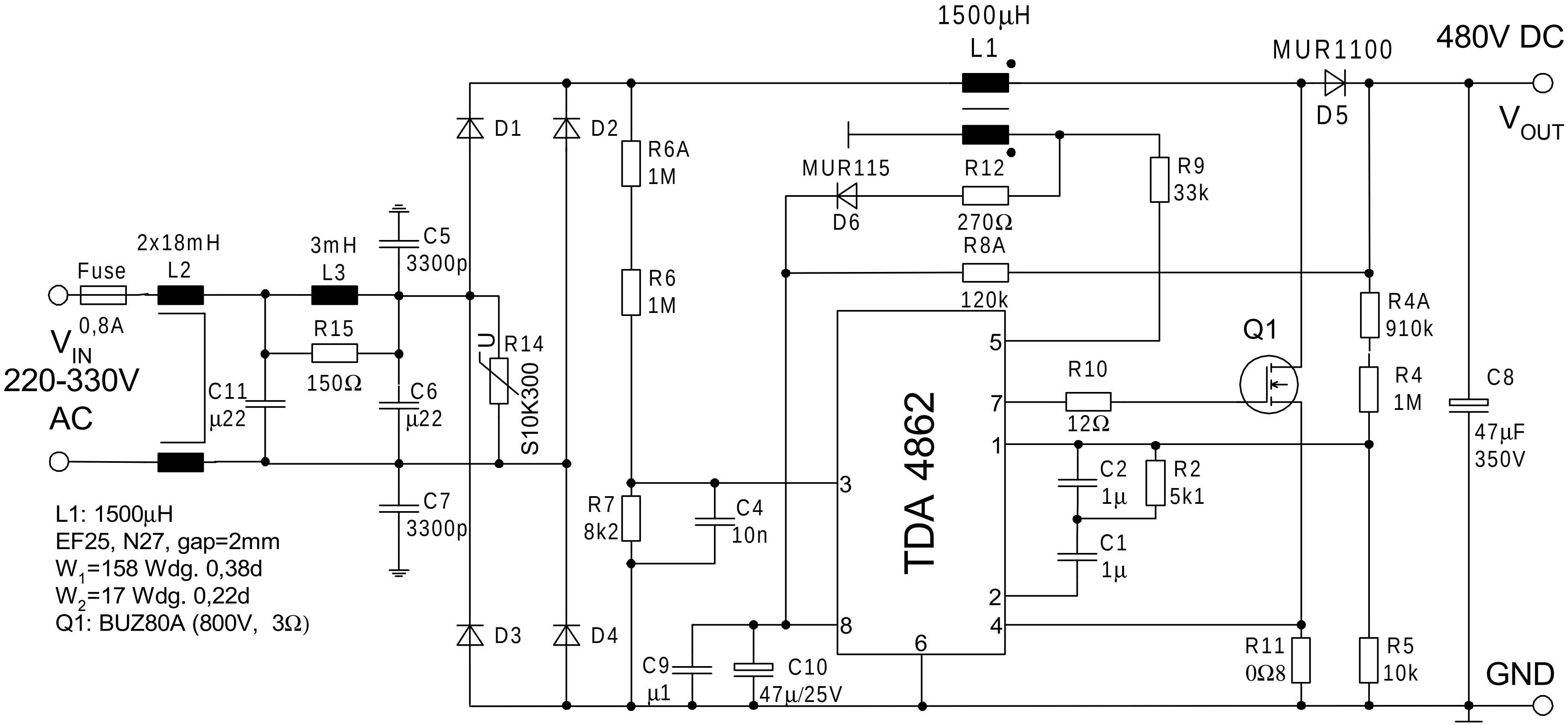

Figure 10: 110W Power Factor Preconverter with TDA 4862 and V_{innom} = 277V Input

Table 3: Measurement of input and output values

277V input for 3 x 35W lamp ballast (Cout = 22µF, L1=1.5mH)										
V_{RMS}	I_{RMS}	P_{IN} real power	PF	THD	V_{OUT}	I_{OUT}	P_{OUT}	V_{OUTAC}	η	I_z (15V) or V_{CC}
220V	0.527A	115.8W	0.999	2.7%	479V	0.229A	110W	35V	0.950	7mA
250V	0.461A	115.1W	0.998	3.8%	479V	0.229A	110W	35V	0.956	3.7mA
277V	0.415A	114.6W	0.996	4.5%	479V	0.229A	110W	35V	0.960	2.1mA
300V	0.382A	114.2W	0.994	5.2%	479V	0.229A	110W	35V	0.963	1.1mA
220V	0.396A	79.3W	0.998	3.2%	479V	0.156A	75W	25V	0.946	7.5mA
277V	0.284A	78.1W	0.991	5.7%	479V	0.156A	75W	25V	0.960	0.7mA
300V	0.263A	77.9W	0.987	6.8%	479V	0.156A	75W	25V	0.963	0.2mA
220V	0.114A	24.3W	0.964	9.5%	479V	0.046A	22W	8V	0.905	0.3mA
277V	0.095A	24.2W	0.916	11.0%	479V	0.046A	22W	8V	0.910	10.7V
300V	0.090A	24.2W	0.889	11.5%	479V	0.046A	22W	8V	0.910	9.9V
220V-300V					479V	0	0	60V		

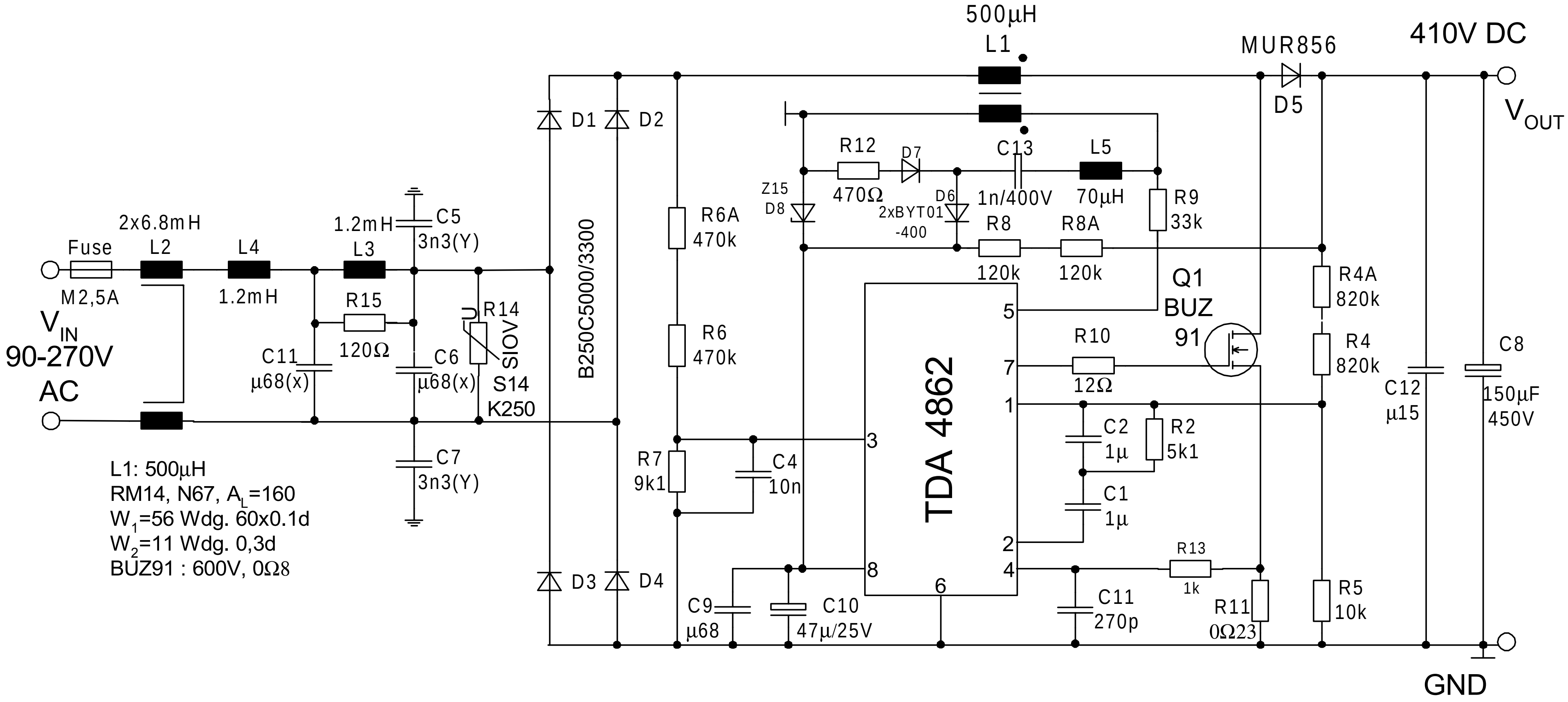

Figure 11: 150W Universal Input Power Factor Preconverter with TDA 4862

Table 4: Measurement of input and output values

90V - 270V / 150W Universal input for SMPS

(Cout = 150μF, L1=500μH)

V_{RMS}	I_{RMS}	P_{IN} real power	PF	THD	V_{OUT}	I_{OUT}	P_{OUT}	V_{OUTAC}	η	I_z(15V) or V_{CC}
90V	1.844A	166.4W	0.998	2.8%	410V	366mA	150W	10Vpp	0.901	0.2mA
120V	1.346A	161.0W	0.999	2.8%	409V	366mA	150W	10Vpp	0.932	1.4mA
180V	0.876A	157.2W	0.996	4.9%	409V	366mA	150W	10Vpp	0.954	4.4mA
230V	0.686A	155.9W	0.987	7.0%	409V	366mA	150W	10Vpp	0.962	6.1mA
270V	0.590A	155.0W	0.973	9.5%	409V	366mA	150W	10Vpp	0.968	6.6mA
90V	0.379A	33.9W	0.994	6.6%	409V	73mA	30W	2Vpp	0.885	5.3mA
120V	0.290A	34.0W	0.981	8.1%	409V	73mA	30W	2Vpp	0.882	8.8mA
180V	0.209A	34.3W	0.911	9.8%	409V	73mA	30W	2Vpp	0.875	12.1mA
230V	0.187A	34.3W	0.798	11.2%	409V	73mA	30W	2Vpp	0.875	9.5mA
270V	0.178A	34.0W	0.708	14.8%	409V	73mA	30W	2Vpp	0.882	4.5mA
180V	0.119A	14.1W	0.66	24.5%	409V	23mA	9.4W	0.8Vpp	0.667	9.6mA
90V-270V					409V	0	0	max. 6Vpp		self-supply

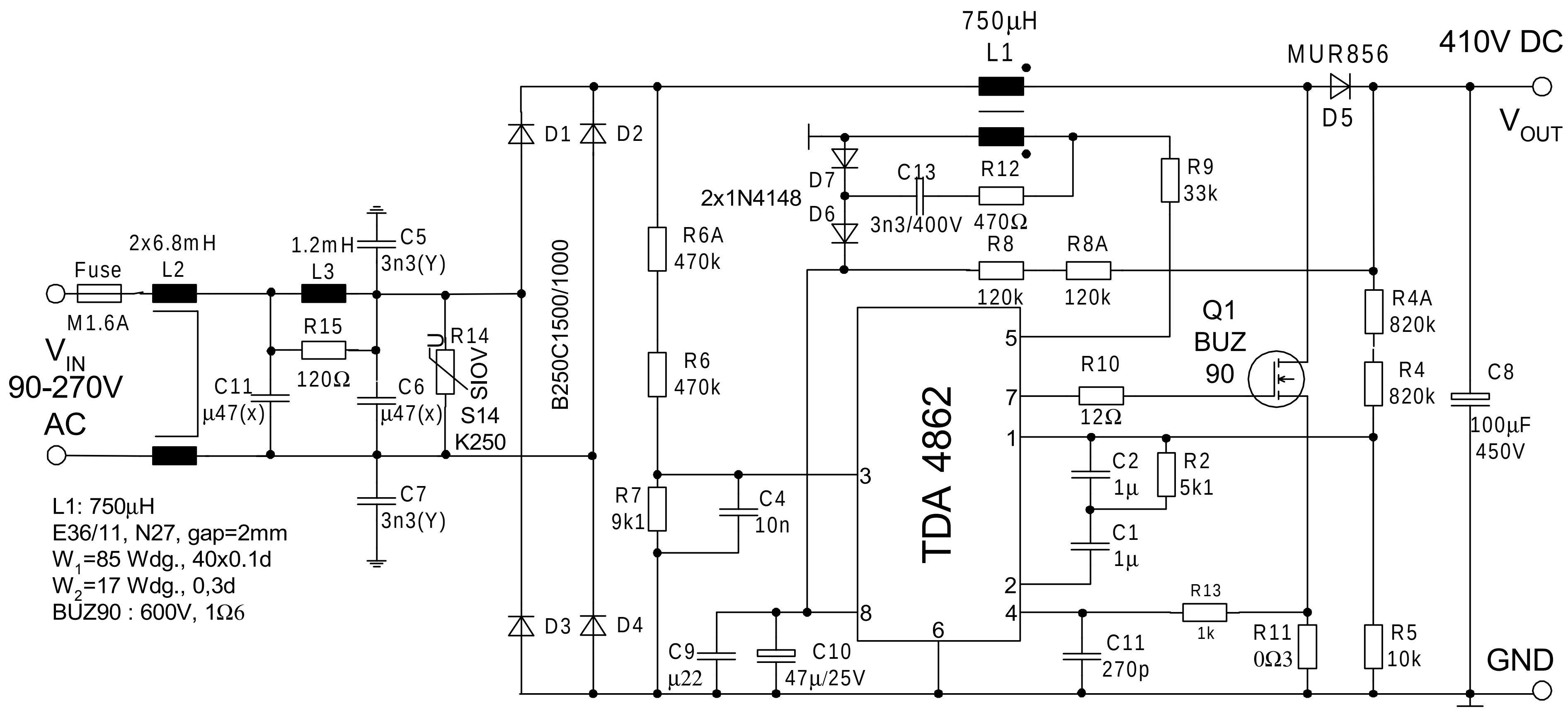

Figure 12: 110W Universal Input Power Factor Preconverter with TDA 4862

Table 5: Measurement of input and output values

90V 270V/110W Universal input for SMPS (Cout = 100µF, L1=750µH)										
V_{RMS}	I_{RMS}	P_{IN} real power	PF	THD	V_{OUT}	I_{OUT}	P_{OUT}	V_{OUTAC}	η	I_z (15V) or V_{CC}
90V	1.355A	122.7W	0.999	2.9%	410V	268mA	110W	11Vpp	0.896	1.2mA
120V	0.984A	118.0W	0.999	3.0%	410V	268mA	110W	11Vpp	0.932	3.2mA
180V	0.643A	115.3W	0.995	5.6%	410V	268mA	110W	11Vpp	0.954	7.0mA
230V	0.505A	115.0W	0.986	8.6%	410V	268mA	110W	11Vpp	0.956	8.2mA
270V	0.434A	114.4W	0.972	11.5%	410V	268mA	110W	11Vpp	0.961	7.3mA
90V	0.280A	25.0W	0.994	7.5%	410V	53.6mA	22W	2Vpp	0.880	5.8mA
120V	0.213A	25.2W	0.984	7.8%	410V	53.6mA	22W	2Vpp	0.873	8.3mA
180V	0.153A	25.4W	0.921	10.2%	410V	53.6mA	22W	2Vpp	0.866	9.6mA
230V	0.132A	25.3W	0.830	9.5%	410V	53.6mA	22W	2Vpp	0.870	7.5mA
270V	0.141A	24.5W	0.646	42%	410V	53.6mA	22W	10Vpp	0.898	0.1mA
90V-270V					410V	0	0	33Vpp		

Summary of used Nomenclature

Physics:

General identifiers:

A .... cross area
b, B ..... magnetic inductance
C .... capacitance
d, D ..... duty cycle
f.......... frequency
i, I........ current
L.......... inductance
N .... number of turns
p, P ..... power
t, T ...... time, time-intervals
v, V ..... voltage
W .... energy
η.......... efficiency

K_1, K_2 .. ferrite core constants

big letters: contant values and time intervals
small letters: time variant values

Special identifiers:

A_L .... inductance factor
$V_{(BR)CES}$ collector-emitter breakdown voltage of IGBT
V_F .... forward voltage of diodes
V_{rrm} ...... maximum reverse voltage of diodes

Components:

C .... capacitor
D .... diode
IC .... integrated circuit
L.......... inductor
R .... resistor
TR....... transformer

Indices:

AC....... alternating current value
DC ...... direct current value
BE....... basis-emitter value
CS....... current sense value
J.......... Junction value
OPTO . optocoupler value
P .... primary side value
Pk .... peak value
R .... reflected from secondary to primary side
S .... secondary side value
Sh .... shunt value
UVLO.. undervoltage lockout value
Z .... zener value

fmin..... value at minimum pulse frequency
i........... running variable
in......... input value
max..... maximum value
min...... minimum value
off .... turn-off value
on .... turn-on value
out ...... output value
p .... pulsed
rip .... ripple value

1,2,3 ... on-going designator

References

[1] **Infineon Technologies AG:** TDA4862 – Power factor and boost converter controller for high power factor and low THD; data sheet; Infineon Technologies AG; Munich; Germany; 07/01.

Revision History		
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Page of actual Rel.	Page of prev. Rel.	Subjects changed since last release
29	29	Deleted
28	28	updated

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