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AT2 9402 E

HL Application Note

Automotive/Transportation/Industrial Electronics

Power Factor Controller TDA 4862 Applications

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Applications:

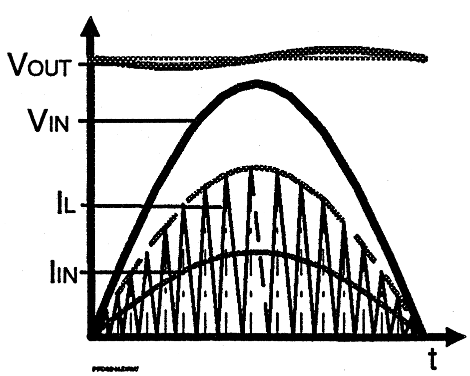
Power Factor Preconverter for lamp ballasts and switched mode power supplies with wide input voltage range.

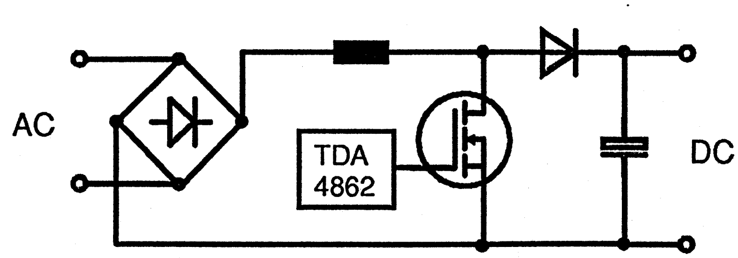
Major features:

- Undervoltage lockout
- One quadrant multiplier
- Zero current detector
- Cycle-by-cycle current limiting
- Internal start-up timer
- 1% bandgap reference
- overvoltage protection
- operating temperature
- -40°C < T_J < 150°C

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.This active harmonic filter limits the harmonic currents resulting from the capacitor pulsed charge currents during rectification. The power factor which describes the ratio between active and apparent power is almost 1. Line voltage fluctuations can be compensated very efficiently.





Technical Description TDA 4862

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, 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 created after smoothing by means of an RFI suppression filter. The RFI suppression filter is to be designed in a way that the valid 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.

Characteristics

Power Supply and Self-Start

An undervoltage lockout with a turn-on threshold of typically 11 V and a turn-off threshold of typically 8.5 V is responsible to make the IC fully functional before the driver output is enabled. In the standby state prior to enabling the driver the IC consumes a current of less than 0.2 mA. A startup timer generates set pulses for the turn-off flip-flop if the driver output takes up L-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.

Driver Output

The driver output has been designed to control power MOSFETs with a current handling 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 handling capability of 100 mA. In the standby state the driver ouput actively controls L-levels using 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 < Tj < 150°C), at its non-inverting input. For the purpose of frequency response compensation a feedback network is inserted between the output (pin 2) and the inverting input (pin 1) of the control amplifier. 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. There is favorable transient response 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 (2.5 V 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 a double mains-frequency AC voltage. The amplitude of the superimposed AC voltage depends on the capacity of the storage capacitor and the load. This superimposed AC voltage, via the control amplifier, causes an undesired modulation of the line current drawn. Therefore a bandwidth of the control amplifier is chosen which is considerably lower than the double mains-frequency. However, this causes the controller to react more slowly to sudden load changes which results in temporary excess voltages and output voltage breakdowns.

Overvoltage Control

If at the boost converter output a higher voltage than the stabilized voltage is generated as a result of voltage transients or load rejection, a current flows from the output voltage divider to the operational amplifier output via the feedback network. This current is measured and in case a threshold of 30 µA is exceeded the multiplier output is controlled to zero potential via a third input. This measure causes the input current to be continously 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 the normal operation state the voltage at the tap of the divider is 2.5 V (= 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 curve form of the feed voltage. In a typical application the rectified and divided supply voltage is applied at 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 curve 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 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 $C_M = 0.65\ V^{-1}$ - an internal amplification factor of the multiplier - is effective, which has the dimension V^{-1} to comply with the following equation. In this way the output voltage of the multiplier V_{QM} which corresponds to the turn-off threshold of the current comparator can be calculated using the following formula:

$$V_{QM} = C_M (V_{pin2} - V_{REF}) V_{pin3}$$

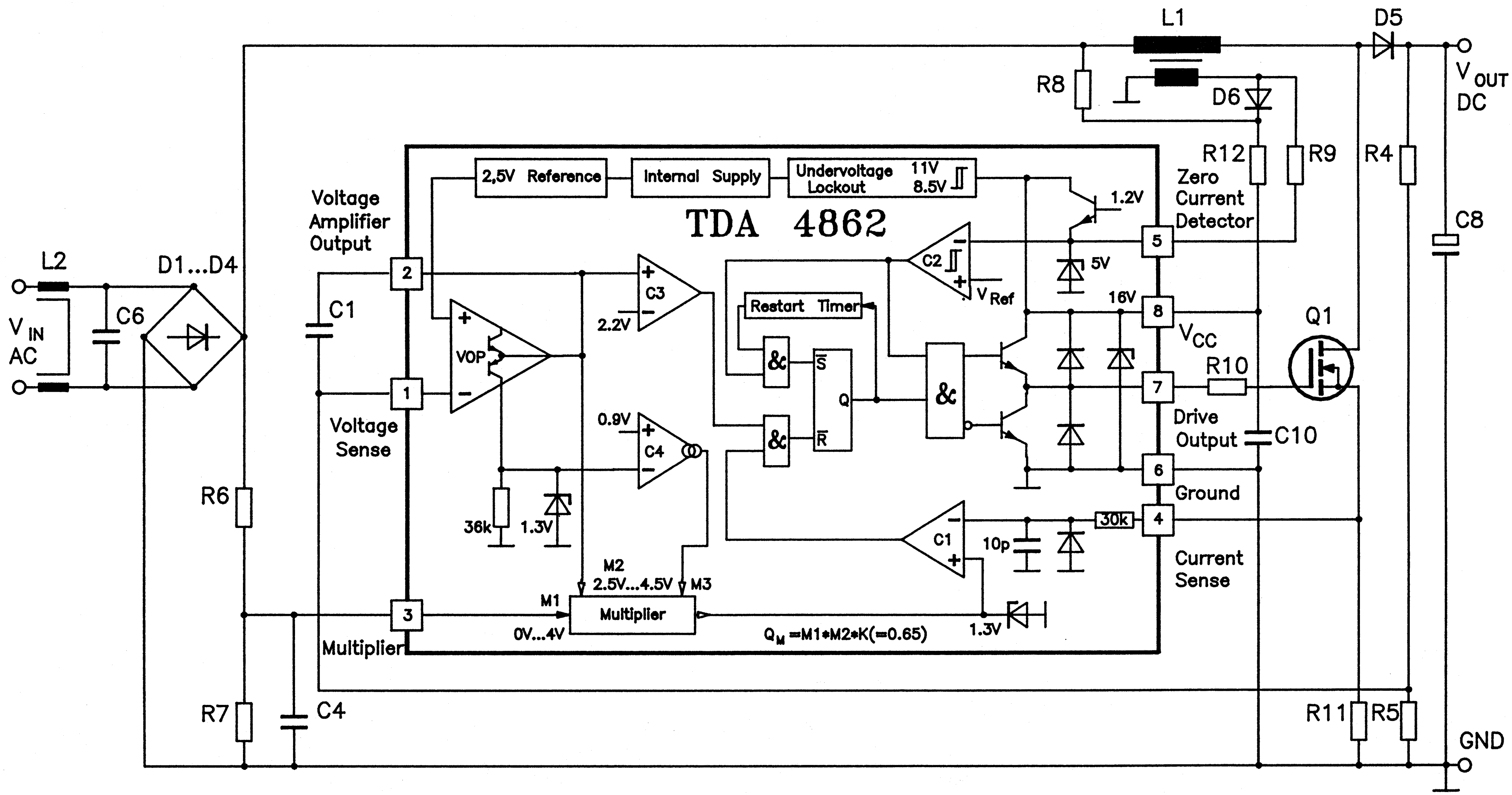
The output voltage of the multiplier is limited to 1.3 V. This measure causes a defined turn-off threshold for current limitation. In this way, dangerous excess currents are avoided which can arise in particular in the case of an expanded input voltage range because the multiplier supplies a higher output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

Current Comparator
The current comparator via its inverting input (pin 4) detects the voltage decline at the shunt which is in the source path of the power MOSFET and should have as low an intrinsic inductance 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 power on and the registration 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 time when the current in the boost converter choke turned zero and then enables the control of a new working cycle. After the current comparator triggered 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 voltage level (V_{OUT}) is available at the drain side terminal of the choke compared with the mains rectifier side terminal (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 main winding, permits the change of polarity of the choke voltage to be registrated 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 accpetance of the "MOSFET blocks" command from the turn-off flip-flop and for setting the flip-flop is 2.5 V (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 (10 K to 47 K). Clamping structures are available in the IC which limit the voltage at the detector 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 controller specifies intermittant operation. In that case a startup generator is activated which supplies a set pulse to the turn-off flip-flop if the driver output stays on L-level longer than 150 µs.



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Typical Application with TDA 4862

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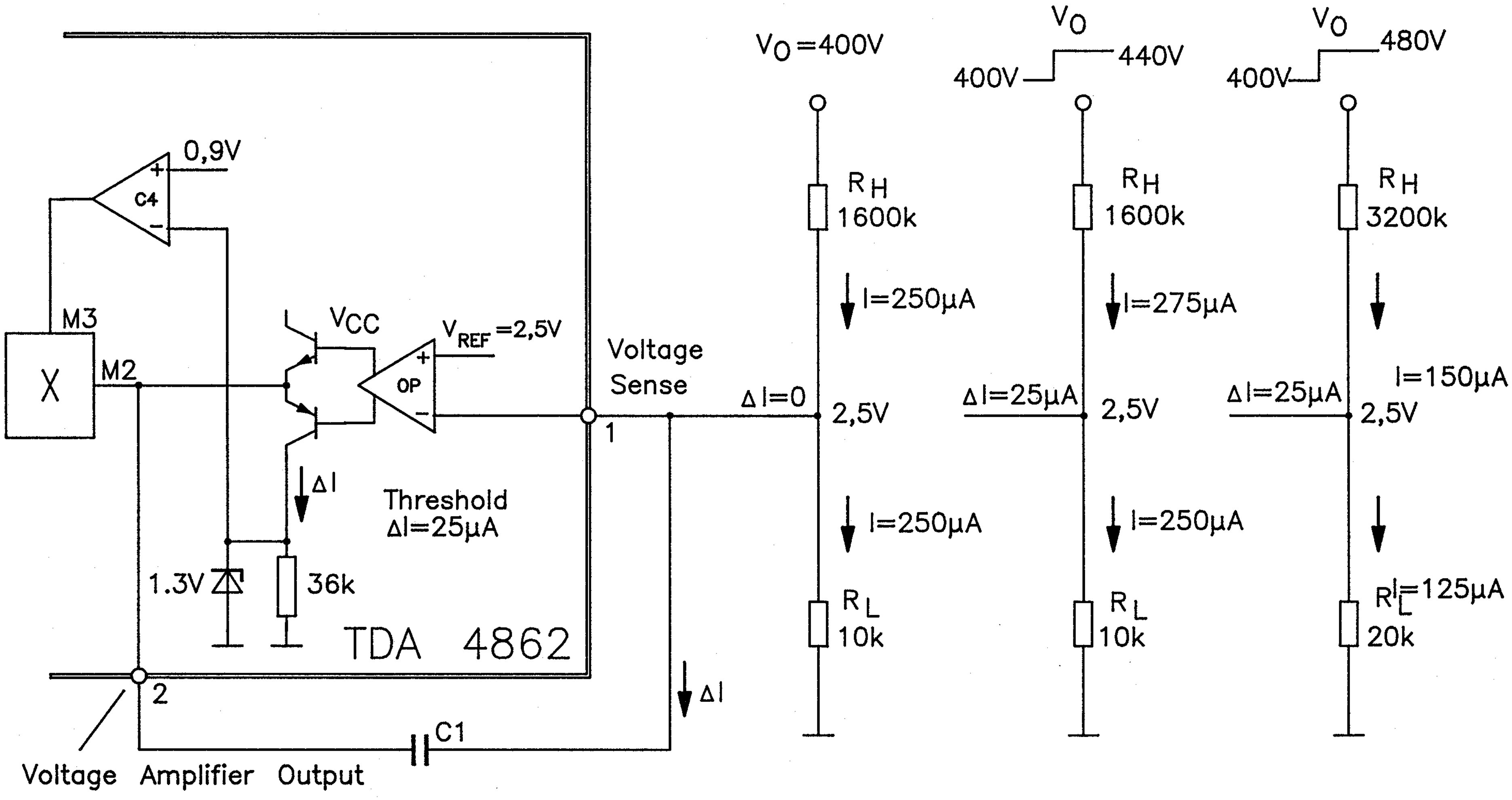

0,9V
C4
M3
X
M2
VCC
VREF =2,5V
OP
Voltage Sense
1
ΔI
Threshold ΔI=25µA
1.3V
36k
TDA 4862
2
ΔI
C1
Voltage Amplifier Output
VO =400V
RH 1600k
I=250µA
ΔI=0
2,5V
I=250µA
RL 10k
VO
400V
440V
RH 1600k
I=275µA
ΔI=25µA
2,5V
I=250µA
RL 10k
VO
400V
480V
RH 3200k
I=150µA
ΔI=25µA
2,5V
I=125µA
RL 20k

Applications of the TDA 4862

The following applications demonstrate the good performance of TDA 4862 controlling a power factor preconverter. The design steps indicate the way to calculate the components values. For example there are three lamp ballast designs and two designs for switched mode mode power supplies (SMPS). Circuit diagrams and measurement results under different operating conditions establish a good basis for evalution.

Beside efficiency, total harmonic distortion (THD) and superimposed output voltage ripple (V_{OUTAC}) there is also a column called I_Z. This is the surplus current of the auxiliary power supply for the IC bypassed through a 15V zener diode. A zener current indicates sufficient power for the IC during the different operating conditions, but it should be low enough to avoid useless losses. There are operating conditions when the zener current reaches zero. Then the actual supply voltage V_{CC} of the TDA 4862 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.

A high efficiency is the result of the discontinuous operation mode avoiding 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 important features of a power factor preconverter. The TDA 4862 includes the control and monitoring functions.

Design steps for a power factor preconverter using TDA4862

Application			2L-Ballast	1L-Ballast	3L-Ballast	SMPS
Input section						
Nominal input voltage	V_{INNOM}		120V AC	230V AC	277V AC	90-270V
Minimum input voltage	V_{INMIN}	$= V_{INNOM}$ -20%	96V AC	184V AC	221V AC	90V AC
Maximum input voltage	V_{INMAX}	$= V_{INNOM}$ +20%	144V AC	276V AC	332V AC	270V AC
Maximum peak input voltage	V_{INPMAX}	$= V_{INMAX} \bullet \sqrt{2}$	204V	390V	470V	382V
Minimum peak input voltage	V_{INPMIN}	$= V_{INMIN} \bullet \sqrt{2}$	136V	260V	313V	127V
Estimated minimum efficiency	η	$= 0.9$				
Output power	P_{OUT}	$= \eta \bullet P_{IN}$	75W	53W	110W	150W
Maximum peak input current	I_{INPMAX}	$= 2 \bullet P_{OUT}/(V_{INPMIN} \bullet \eta)$	1.225A	0.453A	0.781A	2.625A
Maximum HF peak current	$I_{LPMAXHF}$	$= 2 \bullet I_{NPMAX}$	2.45A	0.906A	1.562A	5.25A
Maximum current sense threshold	$V_{ISENSEM}$	$= 1.3V$				
Shunt resistor	R_{11}	$= V_{ISENSEM} / I_{LPMAXHF}$	0.53Ω	1.44Ω	0.83Ω	0.25Ω

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Design steps for a power factor preconverter using TDA4862

Application			2L-Ballast	1L-Ballast	3L-Ballast	SMPS
Output section						
Nominal output voltage	V_{OUT}	(recommended minimum: V_{INPMAX} + 30V)	230V DC	410V DC	480V DC	410V DC
Reference voltage	V_{REF}	= 2.5V				
Regulation current	I_{RVAOUT}	= 25µA				
Overvoltage threshold (selected)		= 10% above V_{OUT}				
Current through R_5	I_{R5}	= I_{RVAOUT} • 100% / 10%	250µA	250µA	250µA	250µA
Output voltage divider	R_5	= V_{REF} / I_{R5}	10K	10K	10K	10K
	R_4	= R_5 (V_{OUT}-V_{REF}) / V_{REF}	910K	1630K	1910K	1630K
Output capacitor C_8						
Capacitance dependend on desired superimposed AC voltage V_{OUTAC}.						
Measured AC output voltage V_{OUTMAC} at C_8 = 1µF and P_{OUT} = 1W.			$13V_{pp}$	$7.9V_{pp}$	$6.4V_{pp}$	$8V_{pp}$
C_8 = P_{OUT} • V_{OUTMAC} / V_{OUTAC}						

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Design steps for a power factor preconverter using TDA4862

Application	2L-Ballast	1L-Ballast	3L-Ballast	SMPS
Multiplier section				
Multiplier input M1 dynamic voltage range $V_{M1R} = 3.8V$				
Multiplier input M2 dynamic voltage range $V_{M2R} = 4.5V - V_{REF} = 2V$				
Multiplier output limitation $V_{QMMAX} = V_{ISENSEM} = 1.3V$				
Multiplier gain K=0.65				
V_{M1}(at $V_{QM} = 1.3V$ and $V_{M2R} = 2V$) = 1.3V/(2V●K) = 1V				
From multiplier characteristics: V_{M1LIM} (at $V_{QM} = 1.3V$ and $V_{M2R} = 2V$) = 1.2V				
Select $V_{M1} = V_{M1LIM} = 1.2V$ at V_{INPMIN} =	136V	260V	313V	127V
Select R_6	1M	2M	2M	940K
$R_7 = R_6 \bullet V_{M1LIM} / (V_{INPMIN} - V_{M1LIM})$	8.89K	9.27K	7.69K	8.95K
Low pass filter $C_4 = 1 / (2 \bullet \pi \bullet R_2 \bullet f)$ {1kHz < f < 3kHz}	10nF	10nF	10nF	10nF
Test input range: V_{M1} (at V_{INPMAX}) < V_{M1} = 3.8V ?	1.80V=OK	1.80V=OK	1.80V=OK	3.62V=OK
otherwise select V_{M1} (at V_{INPMIN}) < V_{M1LIM} = 1.2V				

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Design steps for a power factor preconverter using TDA4862

Inductor of the boost converter

MOSFET on-time $T_{ON} = L \cdot I_{LPHF} / V_{IN}$ off-time $T_{OFF} = L \cdot I_{LPHF} / (V_{OUT} - V_{IN})$

HF inductor peak current $I_{LPHF} = 2 \cdot I_{IN}$ inductivity $= L$

Operating frequency at a definite operating point $f = \frac{1}{T_{ON} + T_{OFF}} = \frac{V_{IN} \cdot (V_{OUT} - V_{IN})}{V_{OUT} \cdot L \cdot I_{LPHF}}$

Recommendation for small input voltage range (e.g. for lamp ballast $V_{IN} = V_{INNOM} \pm 20\%$):

Find L by selecting the operating frequency in a range of 80kHz...110kHz

on nominal input voltage V_{INNOM} and nominal output power

$$L = \frac{V_{INNOM}(V_{OUT} - V_{INNOM})}{V_{OUT} \cdot f \cdot I_{LPHF}} = \frac{V_{INNOM}(V_{OUT} - V_{INNOM}) \cdot \eta \cdot V_{INNOM}}{V_{OUT} \cdot f \cdot 2 \cdot P_{OUT}}$$

Example: $L = \frac{(120V)^2 \cdot (230V - 120V) \cdot 0.9}{230V \cdot 90kHz \cdot 2 \cdot 75W} = 495\mu H$ (e.g. for lamp ballast 75W/120V)

Or by selecting the on-time T_{ON} in the range of 3µs...6µs

$$L = \frac{T_{ON} \cdot V_{INNOM}}{I_{LPHF}} = \frac{T_{ON} \cdot (V_{INNOM})^2 \cdot \eta}{2 \cdot P_{OUT}}$$

Example: $L = \frac{5\mu s \cdot (120V)^2 \cdot 0.9}{2 \cdot 75W} = 432\mu H$ (e.g. for lamp ballast 75W/120V)

Design steps for a power factor preconverter using TDA4862

Inductor of the boost converter

Recommendation for wide input voltage range (e.g. V_{IN} = 90V...270V)

Find L by selecting the operation frequency higher than 25kHz

on maximum peak input voltage and twice of nominal output power (instantaneous value)

and on minimum peak input voltage and twice of nominal output power (instantaneous value)

$$L < \frac{(V_{INPMAX})^2 (V_{OUT} - V_{INPMAX}) \cdot \eta}{V_{OUT} \cdot f \cdot 2 \cdot 2 \cdot P_{OUT}} \qquad L < \frac{(V_{INPMIN})^2 (V_{OUT} - V_{INPMIN}) \cdot \eta}{V_{OUT} \cdot f \cdot 2 \cdot 2 \cdot P_{OUT}}$$

Example: (e.g. for SMPS 150W/90V-270V)

$$L < \frac{(382V)^2 \cdot (410V - 382V) \cdot 0.9}{410V \cdot 25kHz \cdot 2 \cdot 2 \cdot 150W} = 598\mu H \qquad L < \frac{(127V)^2 \cdot (410V - 127V) \cdot 0.9}{410V \cdot 25kHz \cdot 2 \cdot 2 \cdot 150W} = 668\mu H$$

The inductance L should be selected lower than 598µH.

Number of turns: $N = \sqrt{L / A_L}$ $\qquad$ A_L = induction factor

Effective core area: $Ae = \frac{I_{LPHFMAX} \cdot N \cdot A_L}{B_{MAX}} = \frac{2 \cdot I_{INPMAX} \cdot N \cdot A_L}{B_{MAX}}$ $\qquad$ B_{MAX} = max. magnetic flux density

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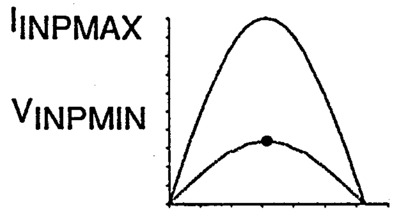
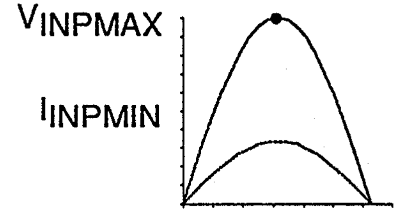
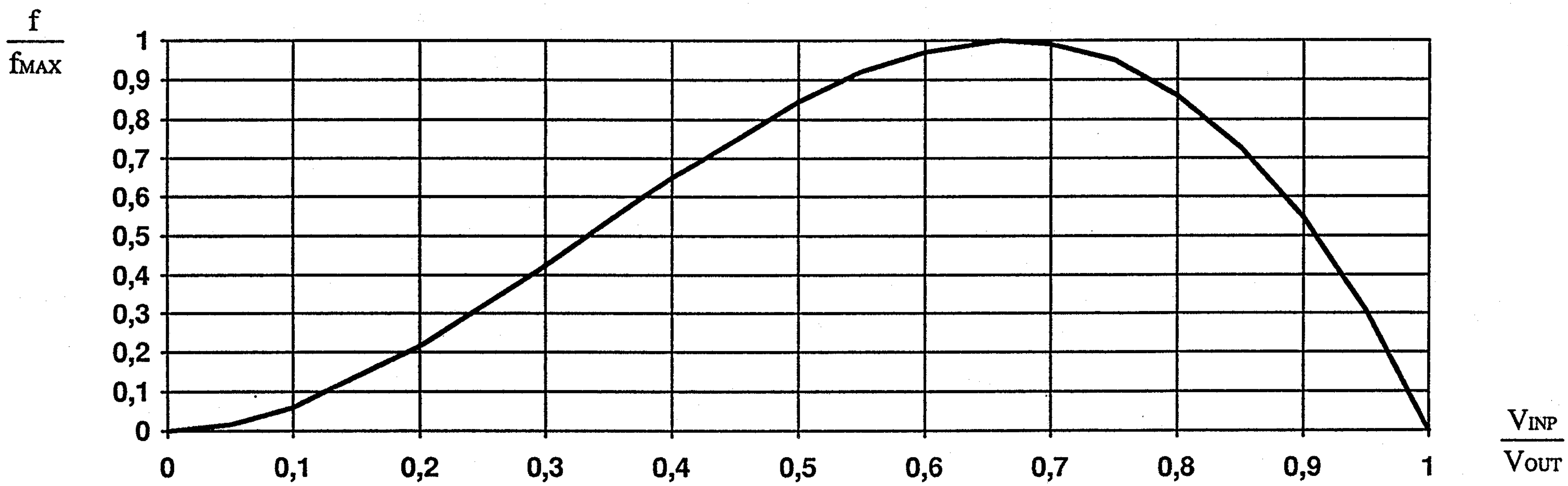
Design steps for a power factor preconverter using TDA4862

Operating frequency f versus peak input voltage V_{INP} at constant output power P_{OUT}

$$f(V_{INP}) = \frac{V_{INP}(V_{OUT} - V_{INP})}{V_{OUT} \cdot L \cdot 2 \cdot I_{INP}} = \frac{V_{INP}(V_{OUT} - V_{INP}) \cdot V_{INP}}{V_{OUT} \cdot L \cdot 2 \cdot 2 P_{IN}} = \frac{V^2_{INP}(V_{OUT} - V_{INP}) \cdot \eta}{V_{OUT} \cdot L \cdot 4 \cdot P_{OUT}}$$







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Design steps for a power factor preconverter using TDA4862

Operating frequency $f(\omega t)$ versus input voltage $V_{IN}(\omega t)$

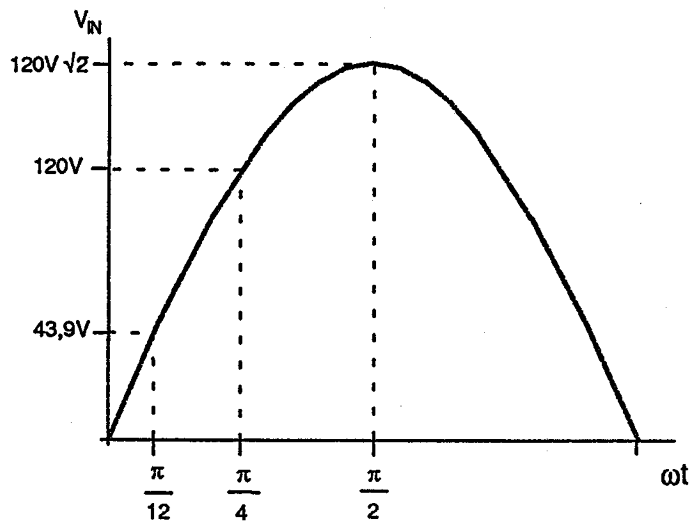
Operating frequency

$$f(\omega t) = \frac{V_{IN} \bullet \sqrt{2} \bullet \sin\omega t \bullet (V_{OUT} - V_{IN} \bullet \sqrt{2} \bullet \sin\omega t)}{V_{OUT} \bullet L \bullet 2 \bullet I_{IN} \bullet \sqrt{2} \bullet \sin\omega t}$$

$$f(\omega t) = \frac{V_{IN}(V_{OUT} - V_{IN} \bullet \sqrt{2} \bullet \sin\omega t)}{V_{OUT} \bullet L \bullet 2 \bullet I_{IN}} = \frac{(V_{IN})^2 \bullet \eta}{V_{OUT} \bullet L \bullet 2 \bullet P_{OUT}} (V_{OUT} - V_{IN} \bullet \sqrt{2} \bullet \sin\omega t)$$

Example

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



$f(\omega t = \frac{\pi}{2}) = 50,3kHz$ $V_{IN}(\omega t = \frac{\pi}{2}) = 169,7V$

$f(\omega t = \frac{\pi}{4}) = 91,8kHz$ $V_{IN}(\omega t = \frac{\pi}{4}) = 120V$

$f(\omega t = \frac{\pi}{6}) = 121,2kHz$ $V_{IN}(\omega t = \frac{\pi}{6}) = 84,9V$

$f(\omega t = \frac{\pi}{12}) = 155,3kHz$ $V_{IN}(\omega t = \frac{\pi}{12}) = 43,9V$

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Design steps for a power factor preconverter using TDA4862

Inductor of the boost converter

Application		Example
Ballast, $V_{INNOM} = 120V$	$L = \frac{(120V)^2 \cdot (230V-120V) \cdot 0.9}{230V \cdot 90kHz \cdot 2 \cdot P_{OUT}} = \frac{34.4mH \cdot W}{P_{OUT}}$	$P_{OUT} = 75W$ $L = 459\mu H$
Ballast, $V_{INNOM} = 230V$	$L = \frac{(230V)^2 \cdot (410V-230V) \cdot 0.9}{410V \cdot 90kHz \cdot 2 \cdot P_{OUT}} = \frac{116mH \cdot W}{P_{OUT}}$	$P_{OUT} = 55W$ $L = 2.1mH$
Ballast, $V_{INNOM} = 277V$	$L = \frac{(277V)^2 \cdot (480V-277V) \cdot 0.9}{480V \cdot 90kHz \cdot 2 \cdot P_{OUT}} = \frac{162mH \cdot W}{P_{OUT}}$	$P_{OUT} = 110W$ $L = 1.47mH$
SMPS, $V_{IN} = 90V\text{-}270V$	$L = \frac{90mH \cdot W}{P_{OUT}}$	$P_{OUT} = 150W$ $L = 600\mu H$

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 max. input voltage V_{INPMAX} and the transformation ratio of the inductor windings have to accomplish the following inequality.

$$(V_{OUT} - V_{INPMAX}) \bullet \frac{W_{ZCD}}{W_{PRIM}} > 2.75V$$

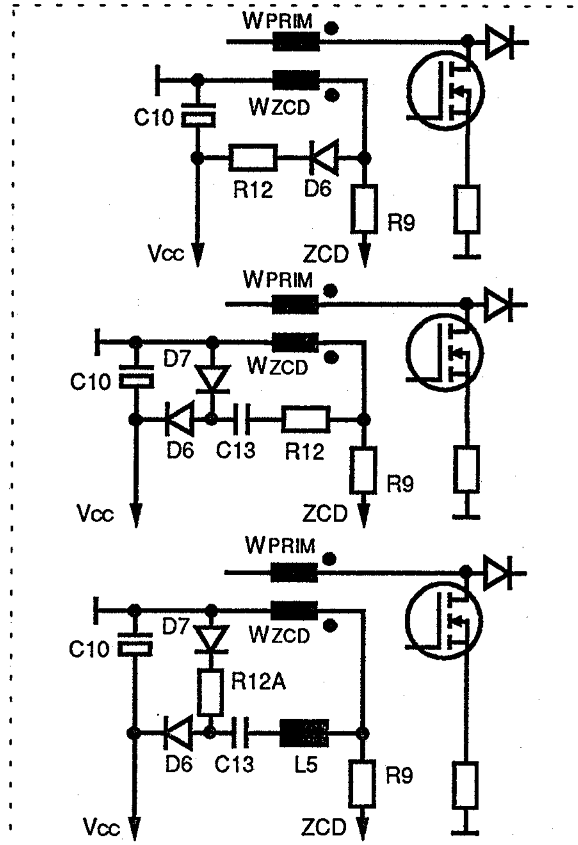
The recommended transformation ratio of $W_{ZCD}/W_{PRIM} = 1/5$ meets a minimal voltage difference of 14V.
If the detector input voltage doesn´t achieve the upper threshold, the IC is operating with the timer frequency.

Auxiliary 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:

$$\frac{W_{ZCD}}{W_{PRIM}} = \frac{V_{ZCD}}{V_{OUT} - V_{INNOM}}$$

V_{ZCD} = 22V...24V
R12 = 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:

$$\frac{W_{ZCD}}{W_{PRIM}} = \frac{V_{ZCD}}{V_{OUT}} \qquad V_{ZCD} \approx 80V,$$

C13 = 3nF...4nF
R12 = 390Ω...470Ω

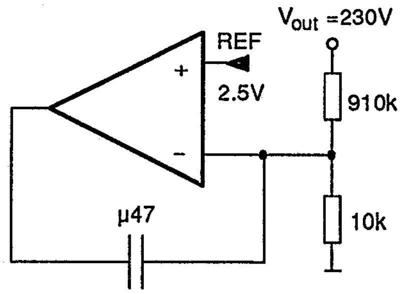
or C13 = 1nF...1.5nF, L5 = 50µH...100µH, R12A designed with C13 and L5 as a low-pass filter of Bessel characteristic.

We achieve the best stability of the output voltage with an error amplifier, which is designed as an integrator.
The low-pass characteristic of the integrator also supports the suppression of the superimprosed AC output voltage.
The gain of the error amplifier should be in the range between 0dB to -10dB at an operating frequency, which is twice of the line frequency.
The boost converter operates like a voltage controlled current source with a capacitor and a load on its output.
The corner frequency of the output configuration is less than 5Hz at nominal load ($f_C = I_{OUT} / (2 \cdot \pi \cdot C_{OUT} \cdot V_{OUT})$).
So the phase margin can reach zero degrees especially at partial load, if a pure integrator is used as an error amplifier.
The result is a poor transient response on load and input voltage changes.

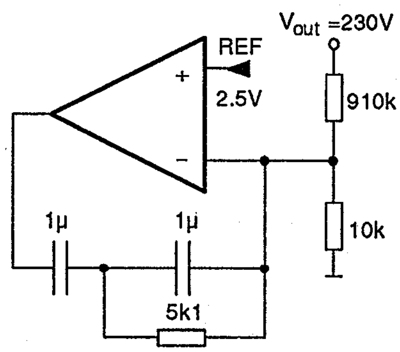
To improve the transient response and the phase margin the error amplifier should have a proportional-integral-low-pass characteristic (PIT1-controller). The proportional section in the frequency range between 10Hz to 33Hz or 16Hz to 33Hz leads to the best results.

Design steps for a power factor preconverter using TDA4862

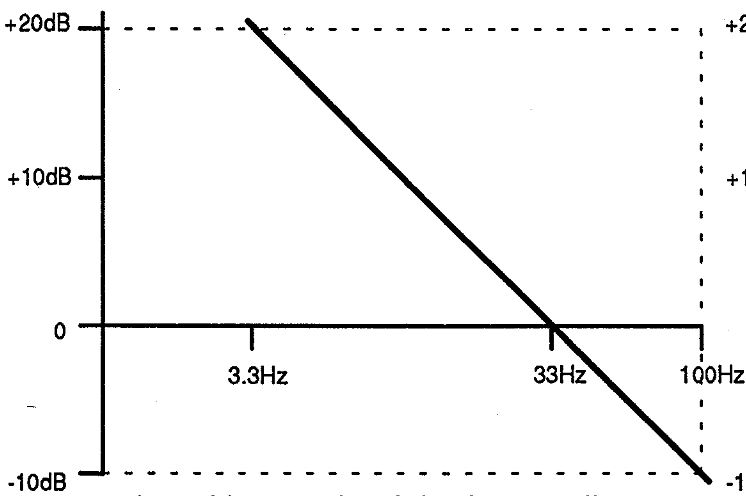
Output voltage error amplifier



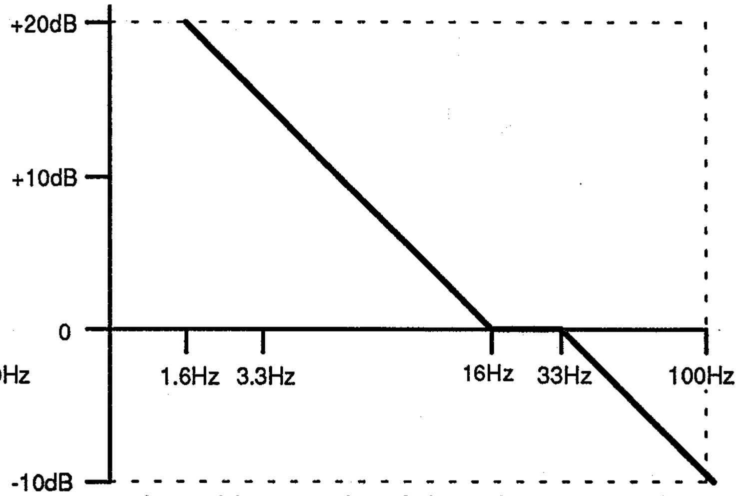
I-controller



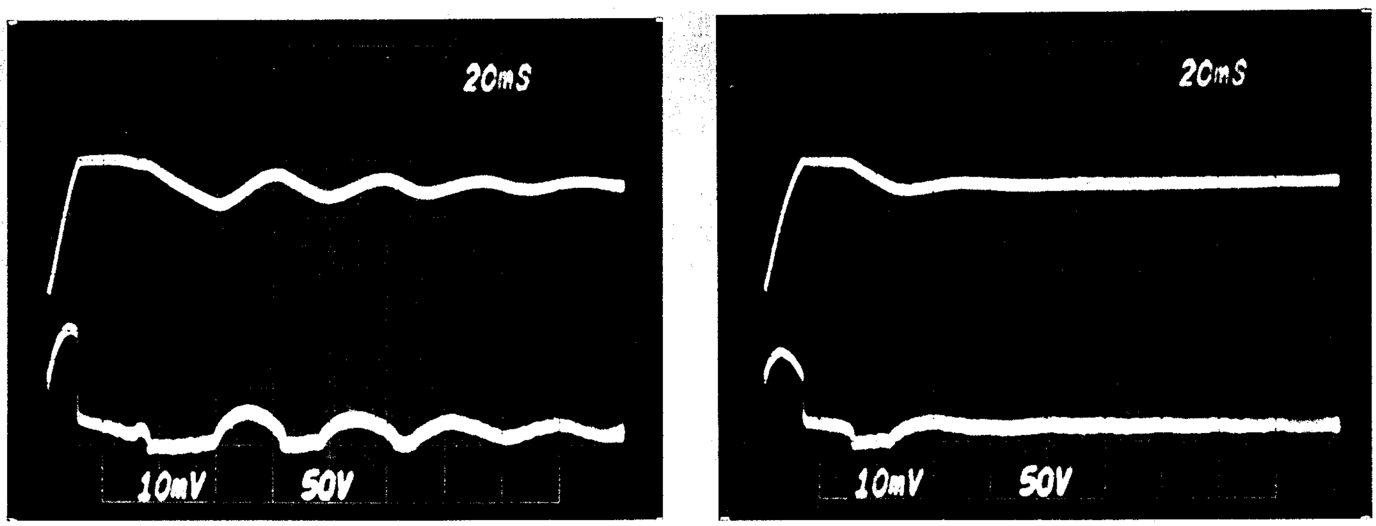
PIT1-controller



closed loop gain of the I-controller



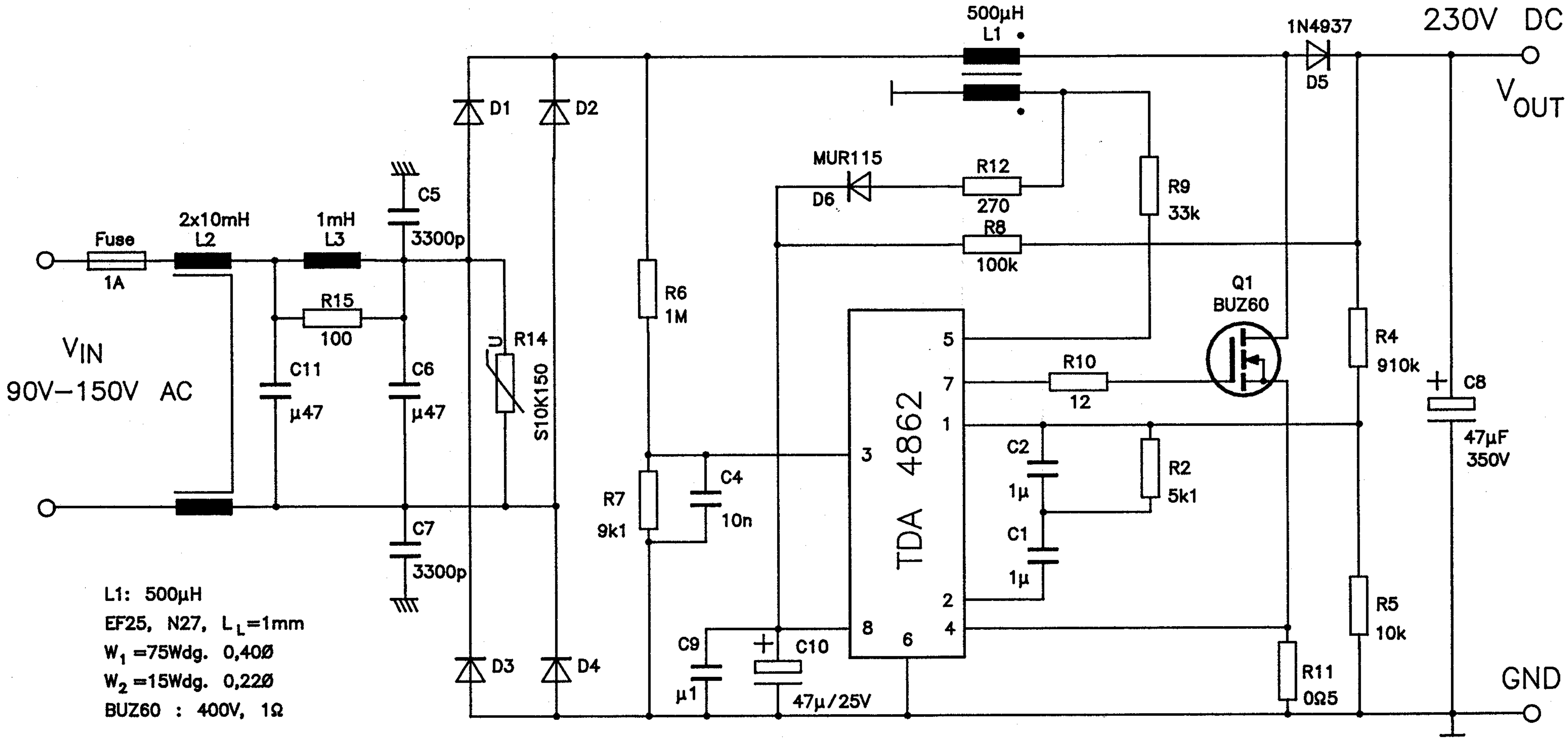
closed loop gain of the PIT1-controller



Transient response of the output voltage during startup at 20% of nominal load

PFC94091.DRW

SIEMENS



75W Power Factor Preconverter with TDA 4862 and 120V Input

4862L212_d

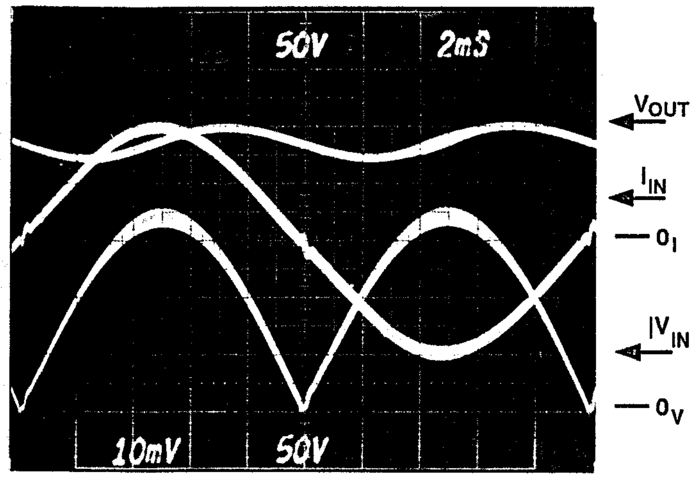
HL AVI AT2 7/94

120V input for 2 x 35W lamp ballast

(C_{OUT} = 47µF , L_1 = 500µH)

V_{RMS}	I_{RMS}	P_{IN} real power	PF	THD	V_{OUT}	I_{OUT}	P_{OUT}	V_{OUTAC} (p-p)	η	Iz(15V) od.VCC
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	0.142A	32.5W	13V	0.923	5.6mA
120V	0.289A	34.45W	0.993	5.0%	229V	0.142A	32.5W	13V	0.943	0.3mA
140V	0.249A	34.25W	0.984	6.5%	229V	0.142A	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					229V	0	0	30V		

Operational behaviour of a 75W Power Factor Preconverter for 2x35W lamp ballast with 120V AC input



nominal load

V_{IN} = 120V AC

$|V_{IN}|$ = 50V/Div

I_{IN} = 0.66A

= 0.5A/Div

V_{OUT} = 229V DC

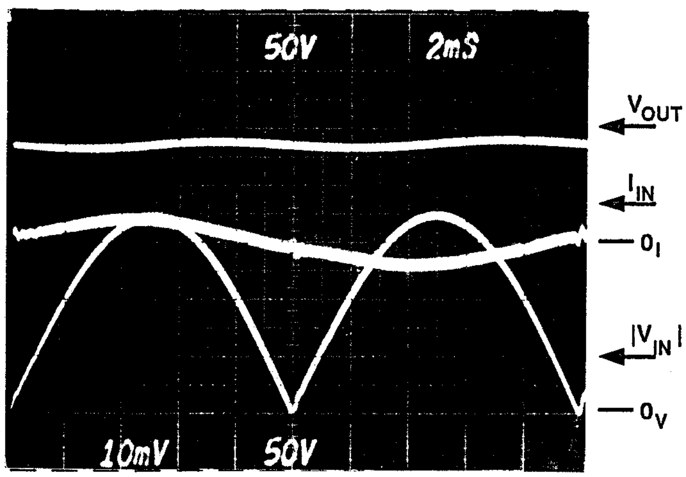
= 50V/Div

T = 2ms/Div

P_{OUT} = 75W

THD (I_{IN}) = 3.2%

PF = 0.999



20% of nominal load

V_{IN} = 120V AC

$|V_{IN}|$ = 50V/Div

I_{IN} = 66mA

= 0.5A/Div

V_{OUT} = 229V DC

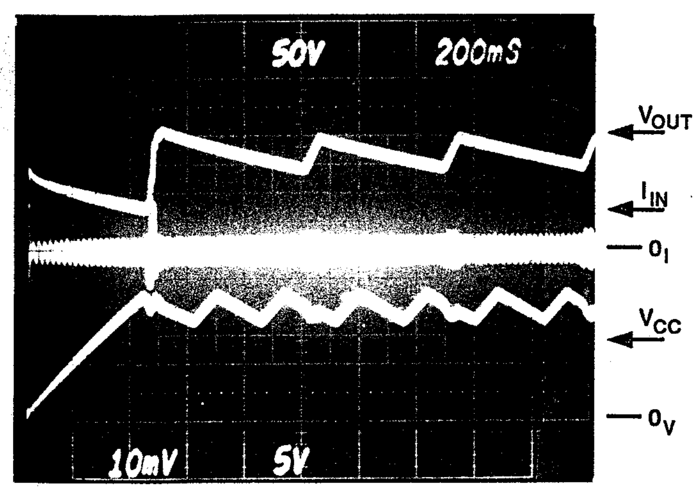
= 50V/Div

T = 2ms/Div

P_{OUT} = 15W

THD (I_{IN}) = 6.8%

PF = 0.965



self start
under no load conditions

V_{IN} = 120V AC

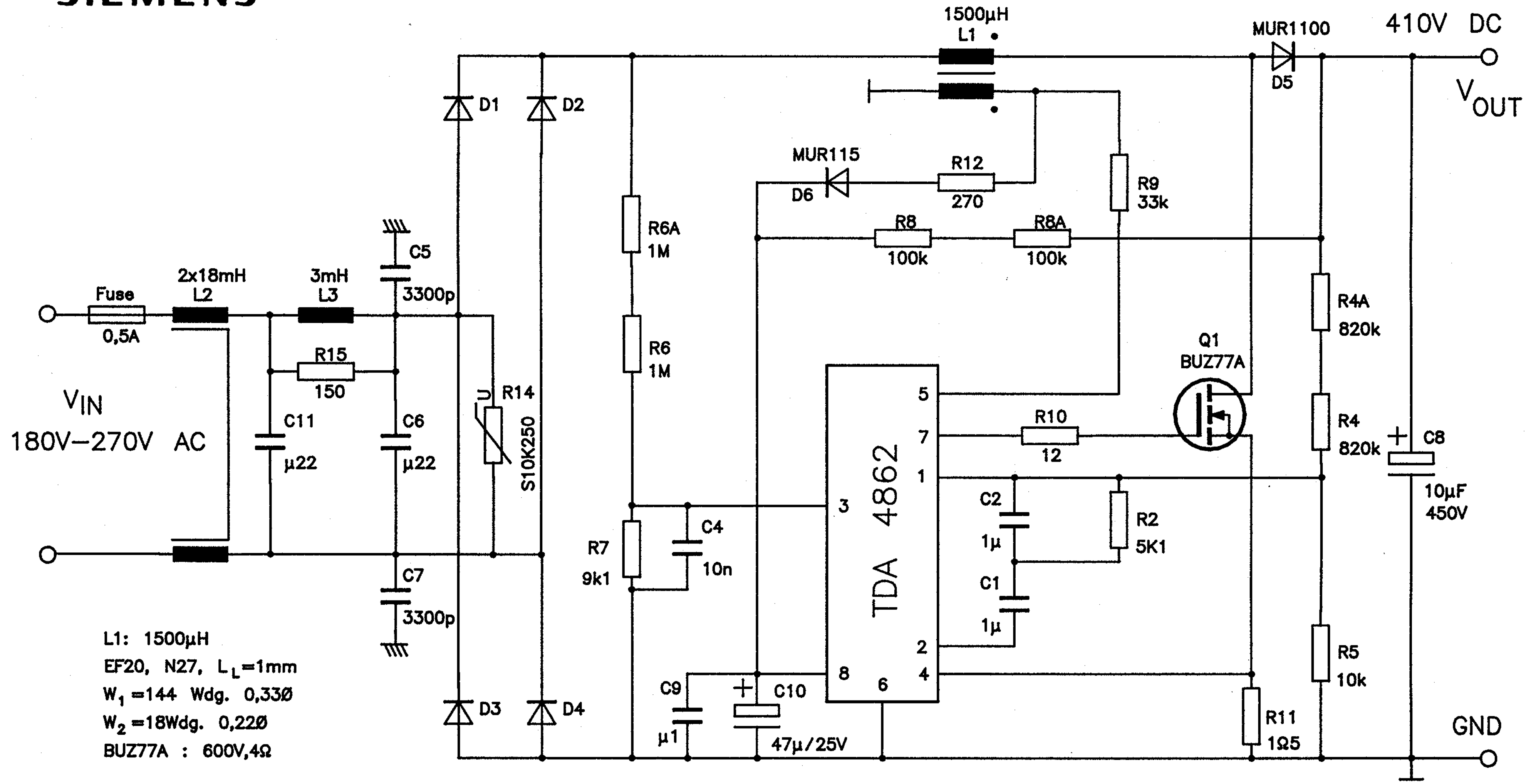
I_{IN} = 0.5A/Div

V_{OUT} = 50V DC

V_{CC} = 5V/Div

T = 200ms/Div

SIEMENS



53W Power Factor Preconverter with TDA 4862 and 230V Input

4862L123_d

HL AVI AT2 7/94

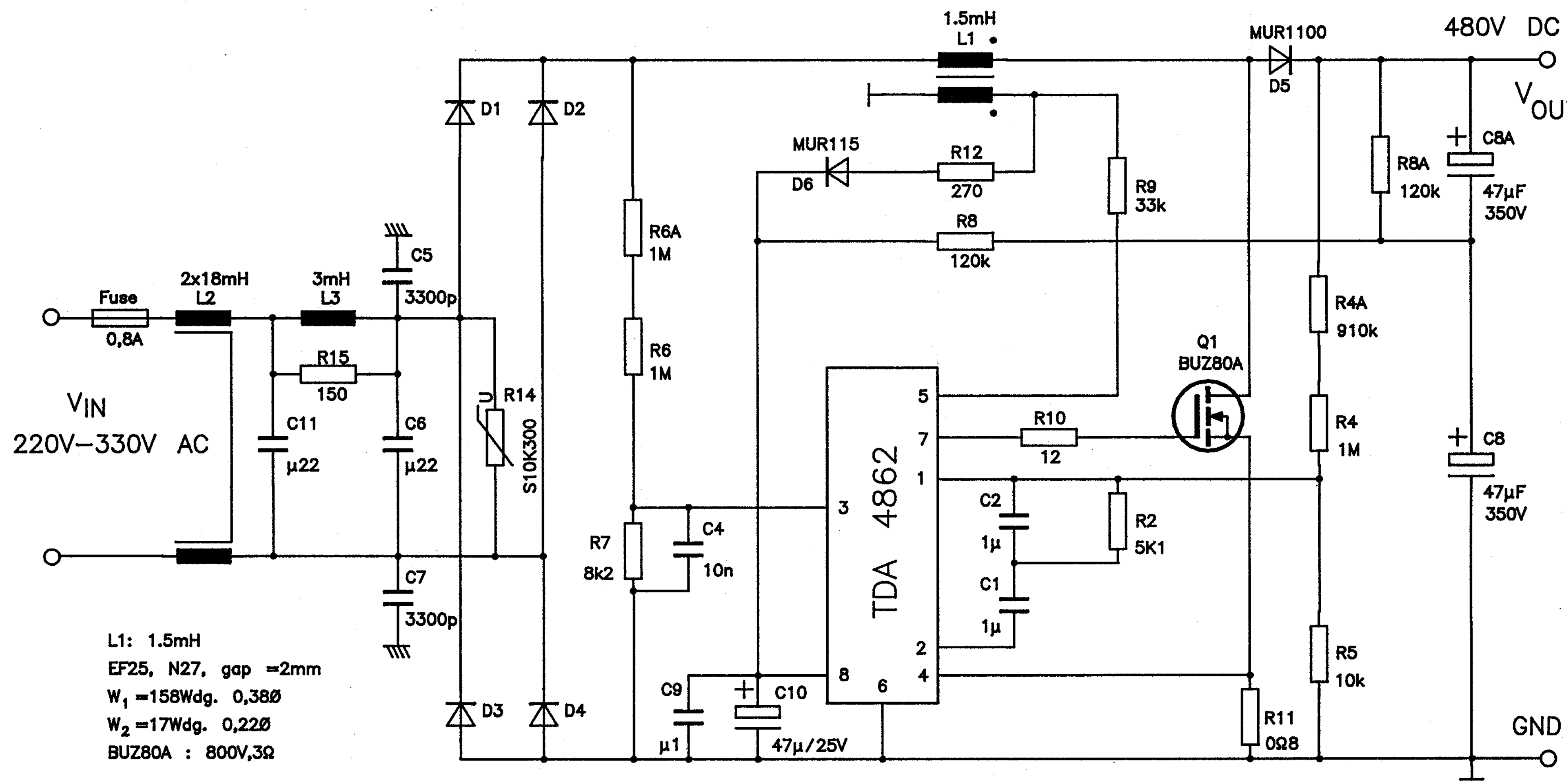
Power Factor Preconverter with TDA 4862

230V Input for 50W lamp ballast

($C_{OUT} = 10\mu F$, $L_1 = 1.5mH$)

V_{RMS}	I_{RMS}	P_{IN} real power	PF	THD	V_{OUT}	I_{OUT}	P_{OUT}	V_{OUTAC} (p-p)	η	I_Z(15V) od.VCC
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
230V	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

SIEMENS



L1: 1.5mH
EF25, N27, gap =2mm
W_1 =158Wdg. 0,38Ø
W_2 =17Wdg. 0,22Ø
BUZ80A : 800V,3Ω

110W Power Factor Preconverter with TDA 4862 and 277V Input

4862L327_d

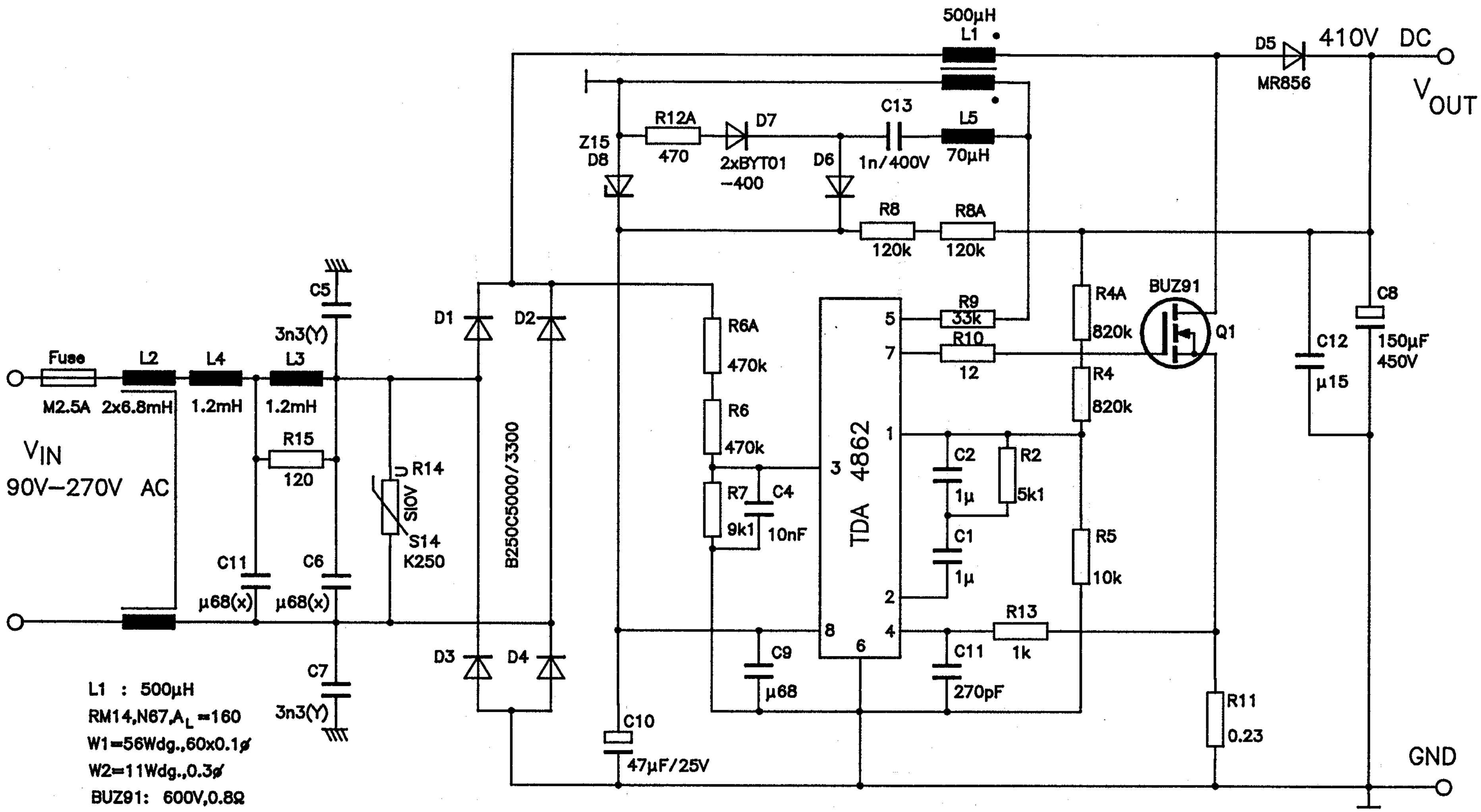
HL AVI AT2 7/94

277V input for 3 x 35W lamp ballast

(C_{OUT} = 22µF , L_1 = 1.5mH)

V_{RMS}	I_{RMS}	P_{IN} real power	PF	THD	V_{OUT}	I_{OUT}	P_{OUT}	V_{OUTAC} (P-P)	η	Iz(15V) od.VCC
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		

SIEMENS



150W Universal Input Power Factor Preconverter with TDA 4862

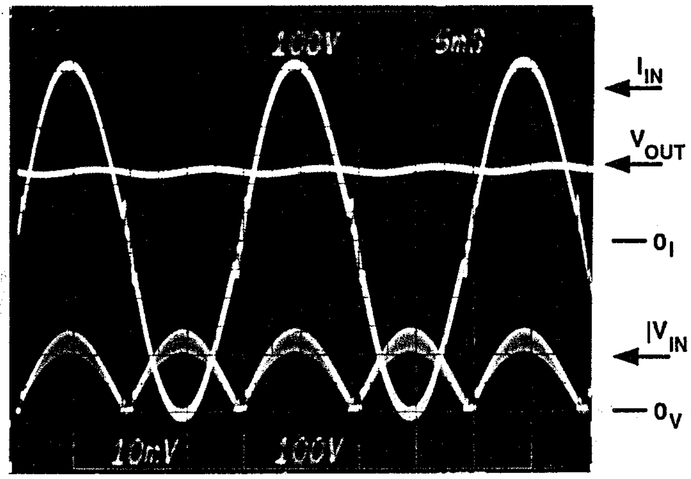
4862W150_d

HL AVI AT 2 10/94

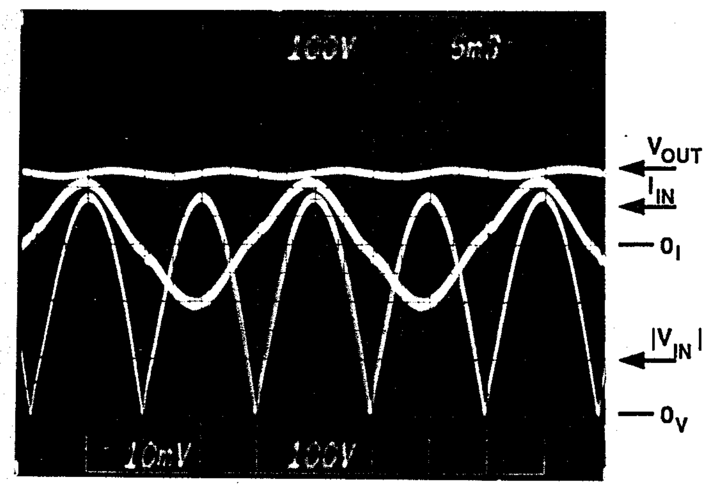
90V-270V / 150W Universal input for SMPS
(C_{OUT} = 150µF , L_1 = 500µH)

V_{RMS}	I_{RMS}	P_{IN} real power	PF	THD	V_{OUT}	I_{OUT}	P_{OUT}	V_{OUTAC} (p-p)	η	I_Z(15V) od.VCC
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

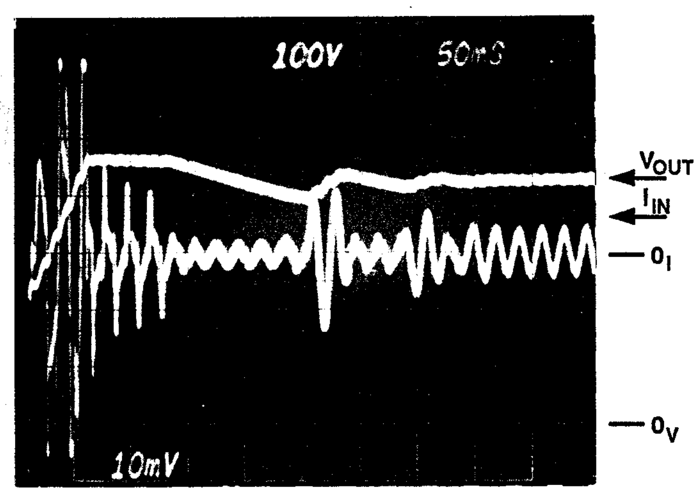
Operational behaviour of a 150W Power Factor Preconverter for SMPS with universal input 90V to 270 V AC



V_{IN} = 90V AC
$|V_{IN}|$ = 100V/Div
I_{IN} = 1.84A
= 1A/Div
V_{OUT} = 410V DC
= 100V/Div
T = 5ms/Div
P_{OUT} = 150W
THD (I_{IN}) = 2.8%
PF = 0.998

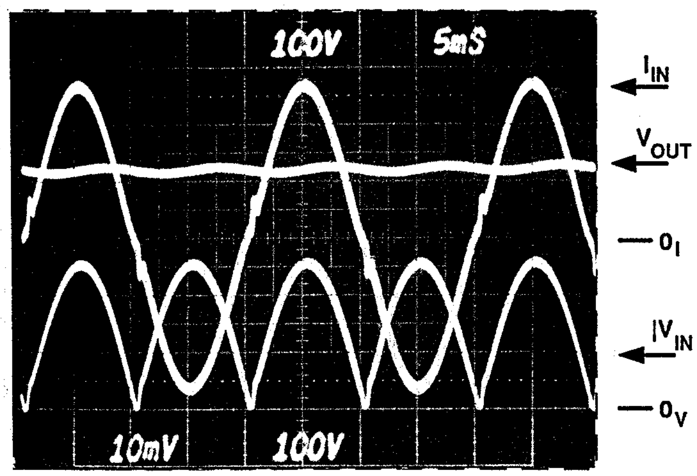


V_{IN} = 270V AC
$|V_{IN}|$ = 100V/Div
I_{IN} = 0.59A
= 1A/Div
V_{OUT} = 409V DC
= 100V/Div
T = 5ms/Div
P_{OUT} = 150W
THD (I_{IN}) = 9.5%
PF = 0.973

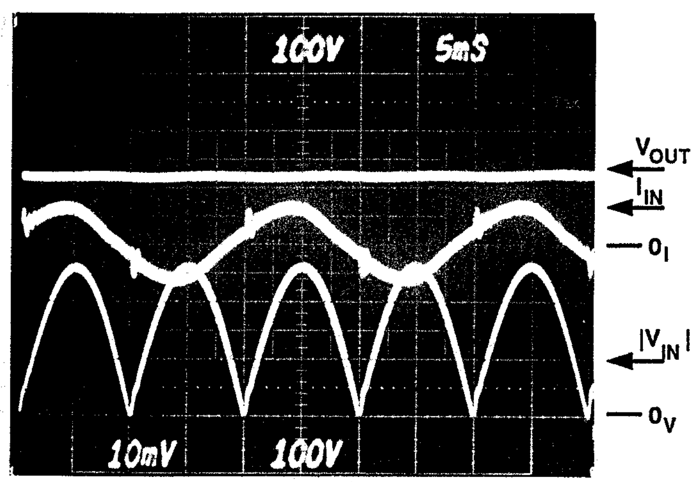


V_{IN} = 180V AC
I_{IN} = 1A/Div
V_{OUT} = 100V/Div
T = 50ms/Div
P_{OUT} = 30W
TDA 4862 power supplied
before input voltage is turned on
→ input current is limited at 3.5A
→ output voltage is limited at 460V

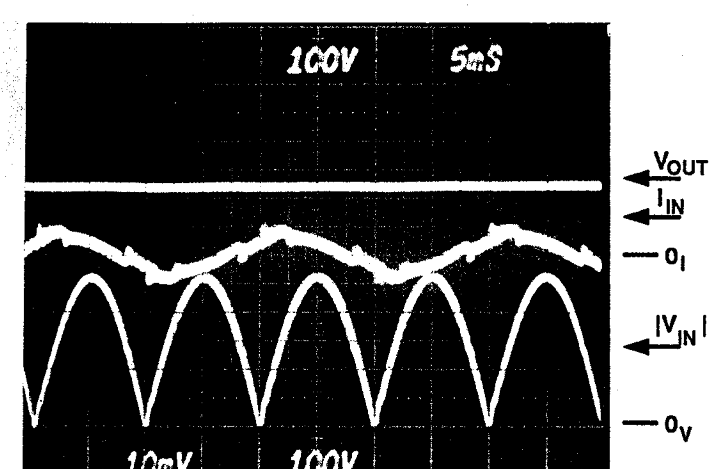
Operational behaviour of a 150W Power Factor Preconverter for SMPS with universal input 90V to 270 V AC



nominal load
V_{IN} = 180V AC
$|V_{IN}|$ = 100V/Div
I_{IN} = 0.876A
= 0.5A/Div
T = 5ms/Div
V_{OUT} = 409V DC
P_{OUT} = 150W
THD (I_{IN}) = 4.9%
PF = 0.996

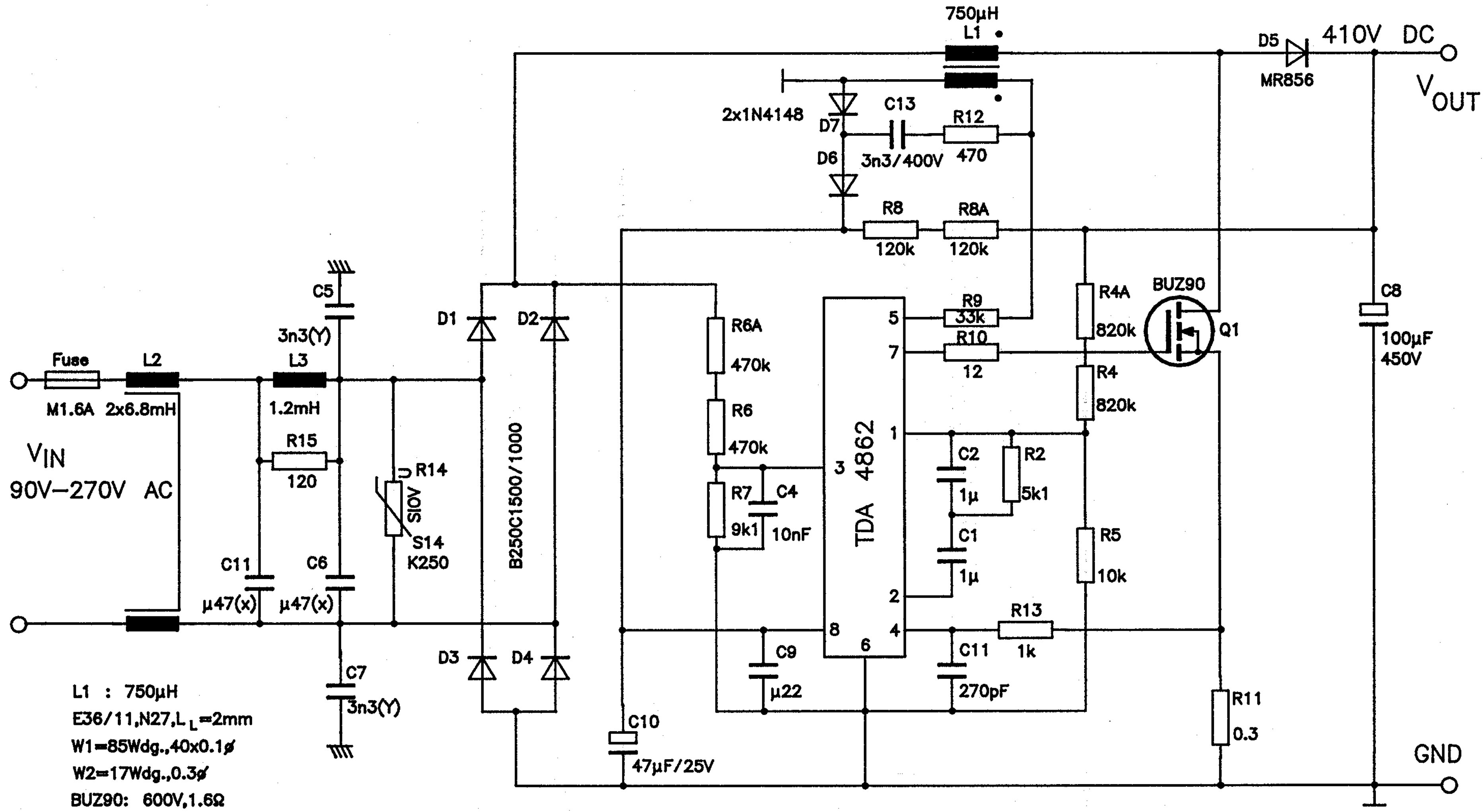


20% of nominal load
V_{IN} = 180V AC
$|V_{IN}|$ = 100V/Div
I_{IN} = 0.209A
= 0.5A/Div
T = 5ms/Div
V_{OUT} = 409V DC
P_{OUT} = 30W
THD (I_{IN}) = 9.8%
PF = 0.911



10% of nominal load
V_{IN} = 180V AC
$|V_{IN}|$ = 100V/Div
I_{IN} = 0.119A
= 0.5A/Div
T = 5ms/Div
V_{OUT} = 409V DC
P_{OUT} = 15W
THD (I_{IN}) = 24.5%
PF = 0.66

SIEMENS



110W Universal Input Power Factor Preconverter with TDA 4862

4862W110_d

HL AVI AT 2 10/94

90V-270V/110W Universal input for SMPS

(C_{OUT} = 100µF , L_1 = 750µH)

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