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Accurate Efficiency Measurements

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2 Scope

The objective of this white paper is to specify a standard; how to determine efficiency η of AC-DC and DC-DC power supply units (PSUs) accurately and comparably. Thermal steady-state is assumed unless stated otherwise. The Second Edition 2018 of this paper is extended with chapters specific to 3-phase systems regarding power factor definition and balanced source conditions.

Deviating measurements result either from random or systematic error. Random errors mainly in the device under test (DUT) are caused by unknown and unpredictable parameters like parts tolerances which can mostly be described by a Gaussian normal distribution. By referring to a reference DUT, random errors are neglected in this guide.

The focus of this paper is on minimizing systematic errors such as:

- Measurement setup (source, sink, wiring, mounting)
- Environmental conditions of the DUT (temperature, altitude)
- Instrument inaccuracy (range, temperature drift)

3 Background

Green marketing issues like “how you can save the world”, an increase in energy costs and last but not least, thermally more sophisticated customer design-ins of power supplies due to miniaturization, lead to increased requirements for high efficiency η , wherein the power losses:

$$P_{loss} = P_{out} \cdot \left(\frac{1}{\eta} - 1 \right)$$

Formula 3.1

dependent on the output power P_{out} is decisive. Since PSU efficiency nowadays rises up to the mid-90s percent, apparently small tolerances in η cause relatively great variance in power losses.

$$\frac{\Delta P_{loss}}{P_{loss}} = - \frac{1}{1-\eta} \cdot \frac{\Delta \eta}{\eta}$$

Formula 3.2

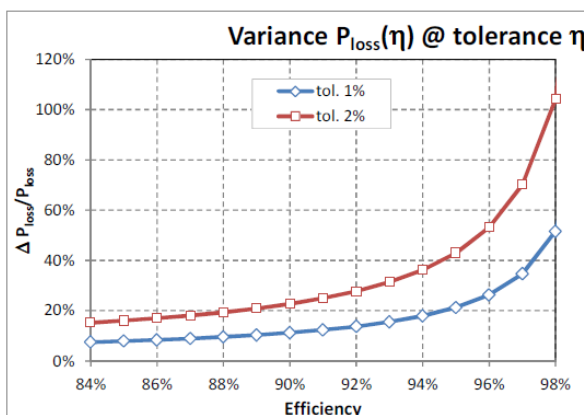


Fig. 3.1

According to Formula 3.2, a small tolerance of 1% in $\eta=95\%$ results in a vast delta in power losses of 20%, which correlates with a temperature rise difference of 20%!

It is obvious that small errors in specifying efficiency may lead to misjudgments, which could cause a severe impact on a customer's thermal design. With higher efficiency, more accurate measurements therefore are vital.

4 Factors affecting efficiency measurements

4.1 Power Factor, Waveform Distortion

4.1.1 Power factor 1-phase; Current harmonic distortion

Factors affecting the input current waveform have to be considered while testing especially in PSUs without active input current control (PFC). The relationships are shown below.

The Power factor λ in general is calculated by:

$$\lambda = \frac{\text{ActivePower}}{\text{ApparentPower}} = \frac{P}{S} = \frac{P_{\text{active}}}{V_{\text{in,rms}} \cdot I_{\text{in,rms}}}$$

Formula 4.1

Assuming a single-phase harmonic input voltage source V_{in} , only the fundamental I_1 of the input current I_{in} can transfer active power by regarding its phase angle Φ_1 . Whereas current harmonics only cause reactive power. Thus, λ can be also described by:

$$\lambda = \frac{V_{\text{in,rms}} \cdot I_{1,\text{rms}} \cdot \cos \varphi_1}{V_{\text{in,rms}} \sqrt{I_{1,\text{rms}}^2 \cdot \cos^2 \varphi_1 + I_{1,\text{rms}}^2 \cdot \sin^2 \varphi_1 + \sum_{n=2}^{\infty} I_{n,\text{rms}}^2}} = g \cdot \cos \varphi_1 = \sqrt{1 - \text{THD}_R^2} \cdot \cos \varphi_1$$

Formula 4.2

V_{in}	sine input voltage
I_{in}	input current
I_1	fundamental (1st harmonic) of input current
Φ_1	phase angle 1st harmonic between V_{in} and I_1
g	fundamental factor of input current
THDR	distortion factor of input current

The apparent power S can also be described as a 3-dimensional vector consisting of the quadrature components active power P , reactive power by fundamental phase shift Q and distortion reactive power D by harmonics >1 .

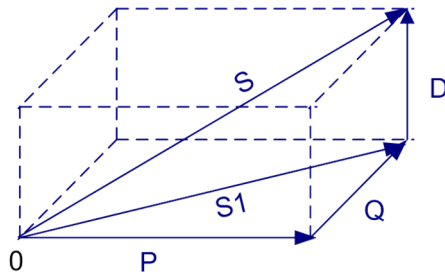


Fig. 4.1

S1 represents the apparent power of the fundamental.

As a consequence of Formula 4.2, by cancelling $V_{in,rms}$ the power factor λ describes the ratio between active and apparent current, which is dependent on the current shape and phase shift. But apparent current causes additional losses in the resistive part R inside the DUT by:

$$P_V = R \cdot I_{rms}^2 = R \cdot \left(\frac{I_{1,rms} \cdot \cos \phi_1}{\lambda} \right)^2 = R \cdot \frac{I_{1,eff}^2}{1 - THD_R^2}$$

Formula 4.3

Crest factor (peak/rms) and current angle are attributes to characterize THD; the higher the crest factor respectively the smaller is the current angle, the more THD and the more losses occur in resistive parts. Thus, factors affecting the input current waveform have to be considered while testing non-linear appliances such as PSUs without active input current control (PFC).

4.1.2 Power factor, 3-phase

Unlike single-phase, a 3-phase system consists of 3 sine voltage sources connected in either Y or Δ and ideally balanced with same amplitude and phase shift of 120° . Since line current is alternating between the 3 sources, the load can't just be described with one Impedance; it should be split in 3 non-linear impedances (in the case without PFC), each causing active and reactive power.

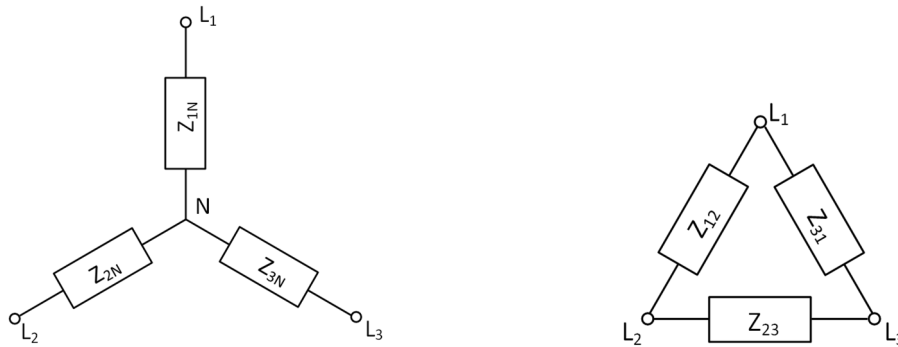


Fig. 4.2 Y-load

Δ -load

In Y-load the current in the impedances equals to the line current and can be measured, but the voltage L-N is unclear since common 3-Phase PSUs don't connect the Neutral. However, when

assuming a balanced 3-Phase load with $Z_{1N} = Z_{2N} = Z_{3N}$, the N-potential can be regarded $\sim 0V$, hence the L-N voltage acts across each impedance. In contrary, the voltage across the impedances in the Δ -load is measurable, but there is no information about the individual currents.

A different approach to get real and apparent power in a 3-phase network is to look at the generated power in each source regardless of the load. Voltage and current for each voltage source can be captured and the consequent real and apparent power derived. The total active power can be calculated by summing all sources of power. But how to calculate the total apparent power S ? Some literature states to proceed analogously by summing P , Q and D of each voltage source.

$$\lambda = \frac{|\bar{P}_1 + \bar{P}_2 + \bar{P}_3|}{|\bar{S}_1 + \bar{S}_2 + \bar{S}_3|} = \frac{P_1 + P_2 + P_3}{|(P_1 + P_2 + P_3) \cdot \bar{p} + (Q_1 + Q_2 + Q_3) \cdot \bar{q} + (D_1 + D_2 + D_3) \cdot \bar{d}|}$$

Formula 4.4

This might be adequate for balanced 3-phase appliances. What if an asymmetric load exists? Does a 2-phase resistive load act with $\lambda = 1$ according to Formula 4. despite unbalanced use of a 3-phase power system?

Standard DIN40110-2 “Quantities in alternating current theory - Part 2: Multi-line circuits” distinguishes more precisely by defining the total power factor for Y-system with:

$$\lambda_{total} = \frac{P_{active}}{S} = \frac{\sum P_{n,active}}{\sqrt{\sum V_{n,rms}^2} \cdot \sqrt{\sum I_{n,rms}^2}} = \frac{P_1 + P_2 + P_3}{\sqrt{(V_{L1-N,rms}^2 + V_{L2-N,rms}^2 + V_{L3-N,rms}^2)} \cdot (I_{L1,rms}^2 + I_{L2,rms}^2 + I_{L3,rms}^2 + I_{N,rms}^2)}$$

Formula 4.5

It is remarkable that Formula 4.5 also takes current in Neutral into account. Hence asymmetric appliances never can reach PF=1!

With identical loss resistance $R_{loss,Line}$ in all Lines (caused e.g. by wires, fuse, terminals, filters) Formula 4.5 can be changed into:

$$\lambda_{total} = \frac{P_{active} \cdot \sqrt{R_{loss,Line}}}{\sqrt{\sum V_{n,rms}^2} \cdot \sqrt{R_{loss,Line}} \cdot \sqrt{\sum I_{n,rms}^2}} = \sqrt{\frac{R_{loss,Line}}{\sum V_{n,rms}^2}} \cdot \frac{P_{active}}{\sqrt{P_{loss,Line,total}}}$$

Formula 4.6

$$P_{loss,Line,total} = \frac{R_{loss,Line}}{\sum V_{n,rms}^2} \cdot \left(\frac{P_{active}}{\lambda_{total}} \right)^2$$

Formula 4.7

For 3-phase Y systems 4.7 gives a stunningly simple correlation between active input power, λ_{total} and the sum of losses in all lines.

It has to be mentioned that apparent power caused by nonlinear impedances differs between Y and Δ systems despite identical line-line voltage at the load. The error is solely caused by different current waveforms causing different distortion reactive power D, whereas both fundamental apparent power S_1 are identical. A plausible explanation from Fig 4.3 is Y-Δ source-conversion increases sinusoidal rms source voltage by the factor of $\sqrt{3}$ whereas non-sinusoidal rms source current is reduced only by $\sqrt{2}$, hence apparent power in Δ-Source is $\sqrt{\frac{3}{2}}$ higher than in Y-sources.

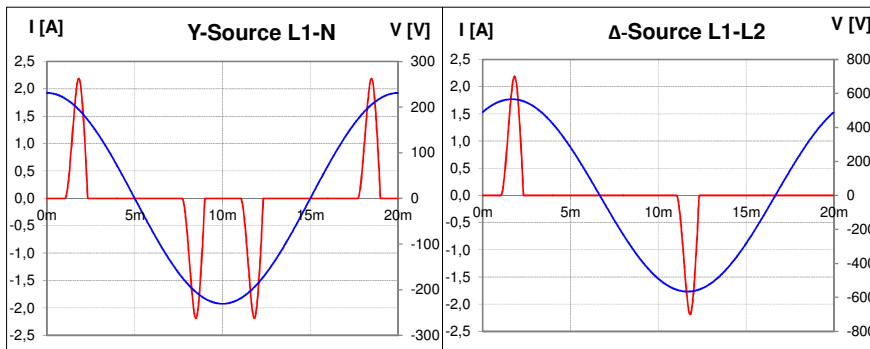


Fig. 4.3 Source voltage and current in Y and Δ with typical non-linear load w/o PFC

4.2 AC-Source

4.2.1 Voltage distortion

With an increasing number of non-linear loads like PCs, Monitors and lamp ballasts, the line voltage is deformed. Flattening of the sine-peak occurs which may reduce the crest factor of the input current especially in non-PFC devices. This lessens the rms value and reduces power losses in the device (Formula 4.3). Furthermore, PFC-stages could get into trouble in regulating the input current to a non-sinusoidal input voltage.

The following measurement examples with different feeding methods at $V_{in}=230V_{ac}$ to a 240W PSU without PFC at nominal load 240W demonstrate quite different results in input current due to deformation of the sine voltage:

AC Power Source:

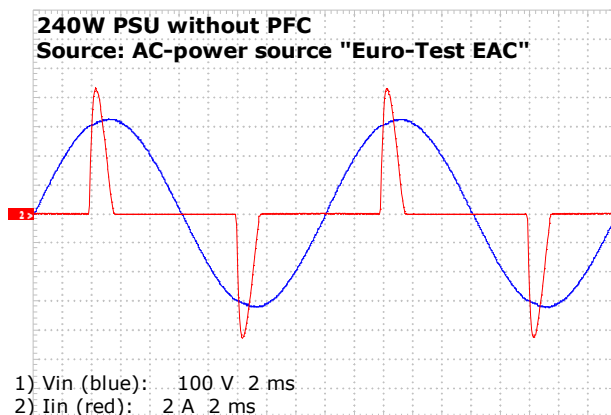


Fig. 4.4

Mains

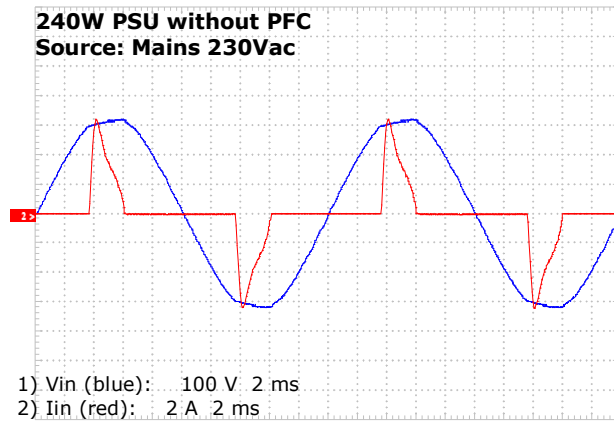


Fig. 4.5

Variable transformer “Thalheimer LTS230/270/6” connected to mains

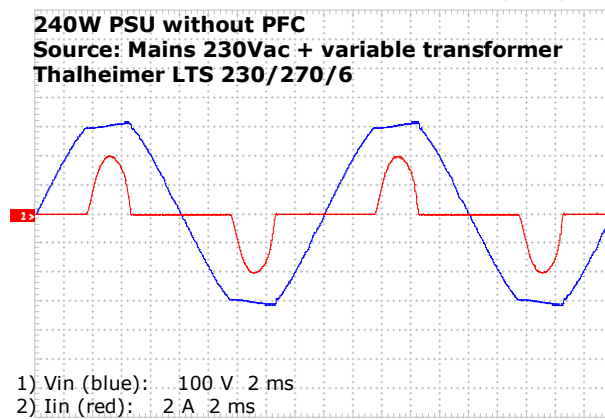


Fig. 4.6

Source	V _{in} [V _{AC}]	I _{in,rms} [A]	S _{in} [VA]	P _{in} [W]	λ	P _{out} [W]	Eta [%]
a. AC Power Source	230,0	2,30	528,94	263,23	0,50	240,0	91,18
b. Mains	229,7	1,93	442,42	262,36	0,59	240,0	91,48
c. Mains + var. Trafo	230,1	1,67	384,35	262,04	0,68	240,0	91,60

Table 4.1 measured with Yokogawa WT3000

According to Table 4.1 measured with Yokogawa WT3000 a deviation of 0.3% in η can be seen between AC power source and mains, and even 0.42% by using a variable transformer. Input current and power factor vary by almost 30%.

The Harmonics Standard IEC61000-3-2 for 1-phase/3-phase mains up to 16A defines the maximum allowed level for current harmonics as well as voltage source distortion in Appendix A.2 paragraph c and d. Mainly the Crest-factor of the Input Voltage at the DUT-terminals should be within 1.40...1.42, the phase of the sine peak within 87...93°.

Conclusion: Only electronic controlled AC power sources complying with IEC61000-3-2 deliver repeatable accuracy and should be used for efficiency measurements.

4.2.2 Line Impedance

The Impedance of the voltage source is especially critical when testing PSUs with no active input current control (power factor correction, PFC). Additional resistance and inductance attenuates the input peak current and widens the current angle reducing input rms current and losses inside the PSU thus making the efficiency seem higher. This includes also the impedance of the wiring. Consequently, line impedance stabilization networks (LISN) e.g. as specified in CISPR 16-1-2 for use with EMI measurements aren't suited because their impedance is too high. Even the use of mains or a variable transformer results in a voltage source too inaccurate in shape and impedance for powering efficiency measurement (see Fig 4.4 and Table 4.1 measured with Yokogawa WT3000). As stated in 4.2.1, solely electronic controlled AC-power sources complying with IEC61000-3-2 deliver repeatable accuracy. But make sure there is no clipping as the peak current may be limited by the source! Observing the input voltage with a 'scope is recommended to ensure a clean sine-wave.

Some Simulation examples of a 240W PSU without PFC with different source impedances @230Vac/50Hz:

Source Impedance 10mΩ/10μH

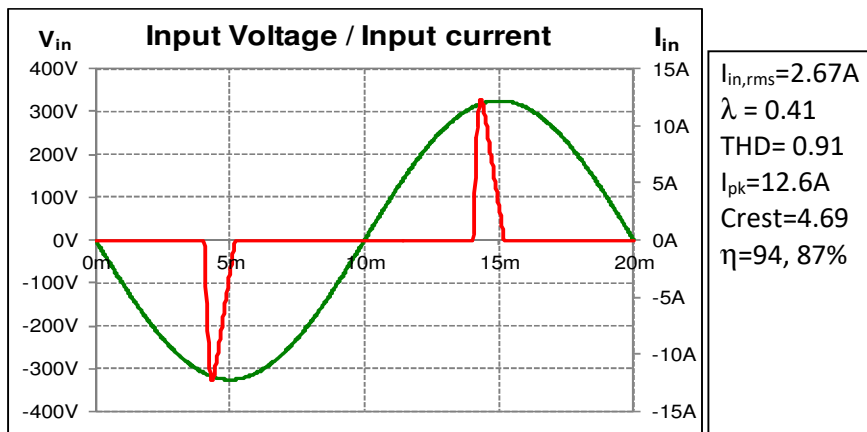


Fig. 4.7

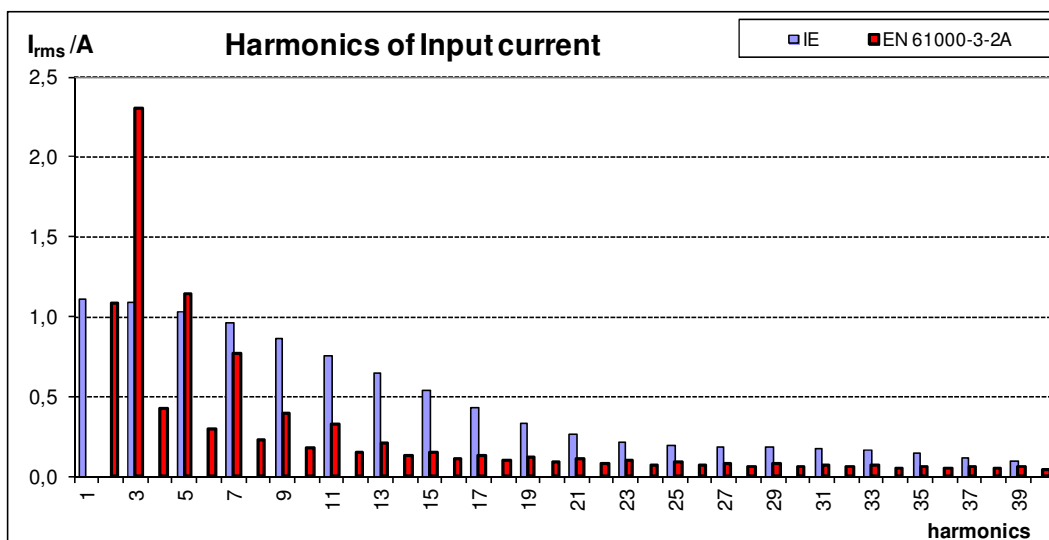


Fig. 4.8

Source impedance 1Ω

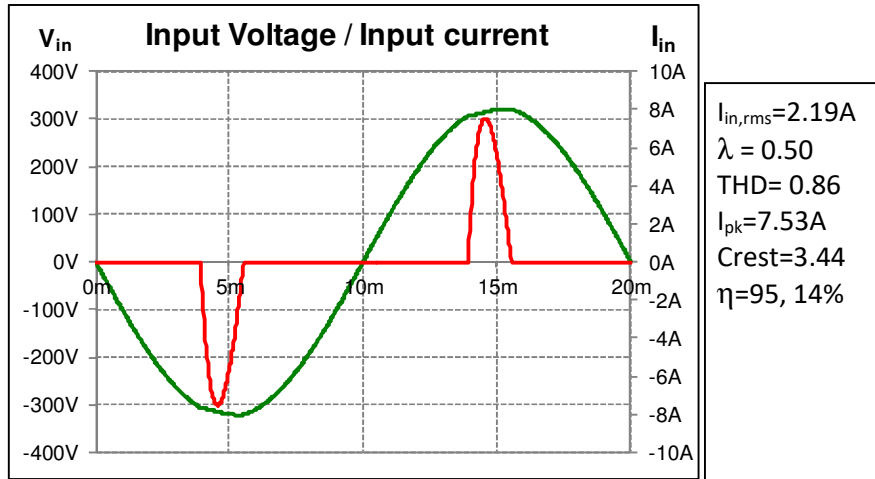


Fig. 4.9

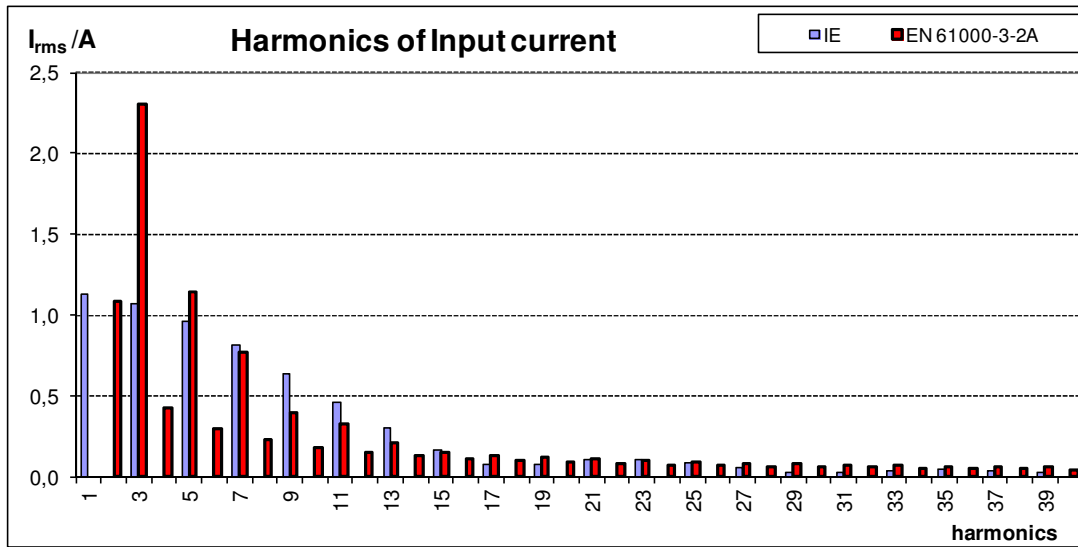


Fig. 4.10

Source impedance 1Ω /1mH

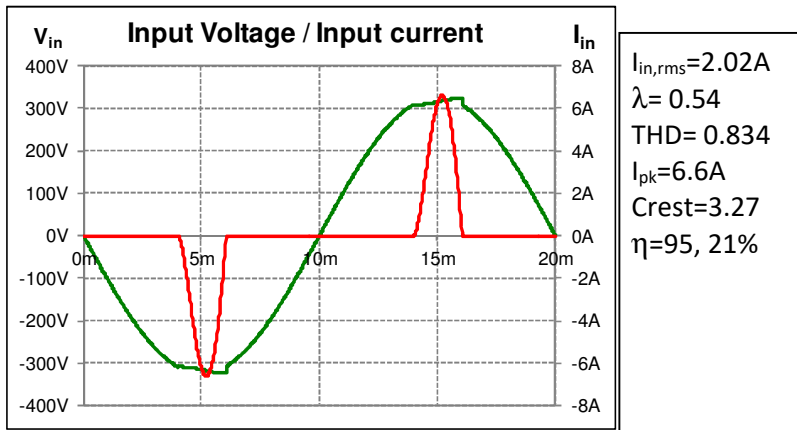


Fig.4.11

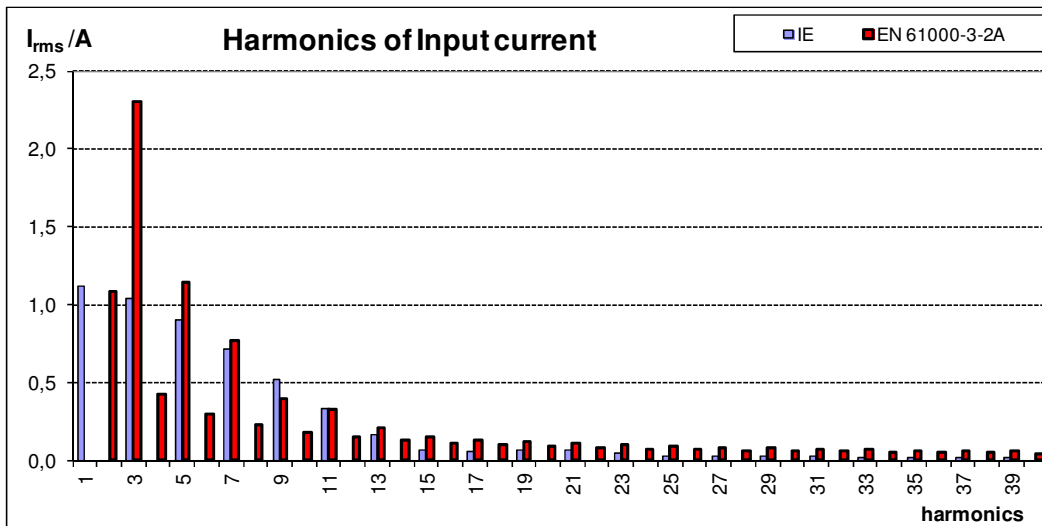


Fig. 4.12

The simulation demonstrates that the input impedance has a significant influence on harmonics; an additional resistance of 1 Ohm minimizes peak current almost by a factor of 2 reducing input rms current by 32% and increasing η by 0.34%.

Conclusion: In order to keep the λ error <0.01 and the losses in the input wiring <0.1% of the active input power, the total input wiring resistance should not exceed:

$$R_{wiring} < \frac{\lambda^2}{1000} \cdot \frac{V_{in,rms}^2}{P_{in,active}}$$

Formula 4.8

4.2.3 Frequency

Worldwide, the public network frequency varies between 50Hz and 60Hz. Special networks like railway (16.7Hz) and aircraft systems (400Hz) have higher deviation. Regarding again PSUs without PFC, it can be said, with lower source frequency, the voltage ripple on the bulk capacitors increases, expanding the current angle and thus minimizing harmonics and rms value. Simulation examples for various source frequencies with a 240W PSU without PFC @230Vac Source Impedance 10mΩ/10μH

$F = 47\text{Hz}$

$I_{in,rms} = 2.672\text{A}$ $\lambda = 0.412$ $\text{THD} = 0.908$
 $I_{pk} = 12.526\text{A}$ $\text{Crest} = 4.688$ $\eta = 94.89\%$

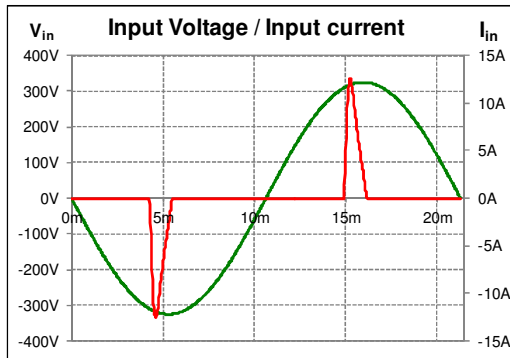


Fig. 4.13

$f = 63\text{Hz}$

$I_{in,rms} = 2.761\text{A}$ $\lambda = 0.399$ $\text{THD} = 0.916$
 $I_{pk} = 12.9\text{A}$ $\text{Crest} = 4.672$ $\eta = 94.82\%$

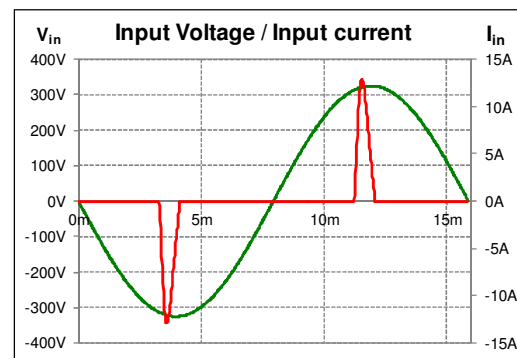


Fig. 4.14

$F = 16.7\text{Hz}$

$I_{in,rms} = 2.19\text{A}$ $\lambda = 0.5$ $\text{THD} = 0.845$
 $I_{pk} = 8.76\text{A}$ $\text{Crest} = 4.0$ $\eta = 95.2\%$

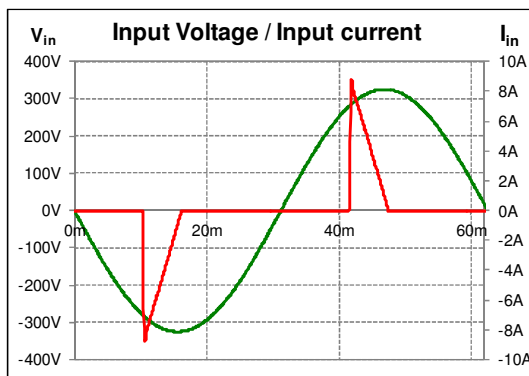


Fig. 4.15

$f = 400\text{Hz}$

$I_{in,rms} = 2.39\text{A}$ $\lambda = 0.459$ $\text{THD} = 0.888$
 $I_{pk} = 9.43\text{A}$ $\text{Crest} = 3.93$ $\eta = 94.99\%$

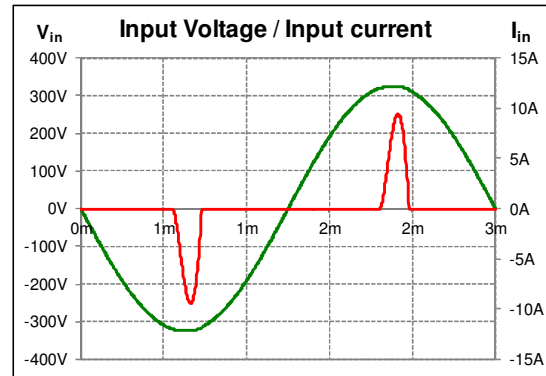


Fig. 4.16

The influence of frequency in the range of 47-63Hz is slightly noticeable in the current peak (3%), but almost negligible in efficiency. Hence, it's accurate enough to specify efficiency in the range of 50-60Hz. Significantly less current distortion occurs at 16.7Hz resulting in an increase in efficiency by 0.3%. There is amazingly small THD at 400Hz. Inductance in the input circuit like stray inductance of a Common mode choke smooths the current.

Efficiency comparison measurements between a 240W active-PFC PSU and a non-PFC version both fed by an AC power source 230Vac at nominal output power confirm insignificant influence of the line (net) frequency. (Power Analyser: Yokogawa W3000, AC-Source power source).

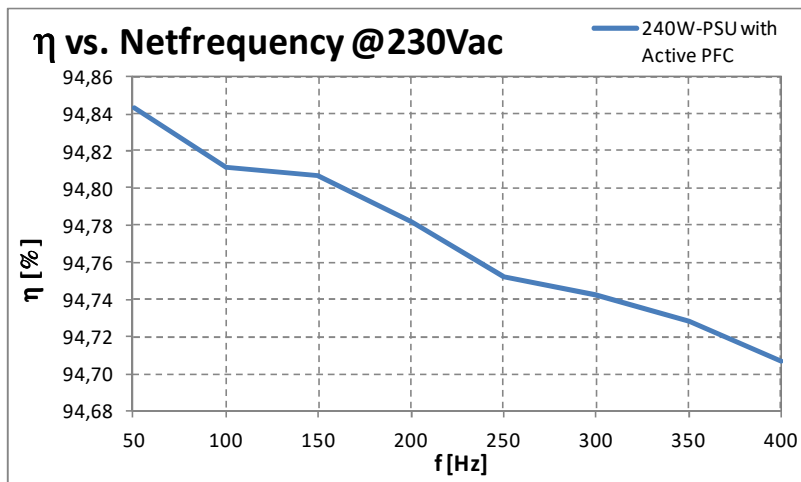


Fig 4.17

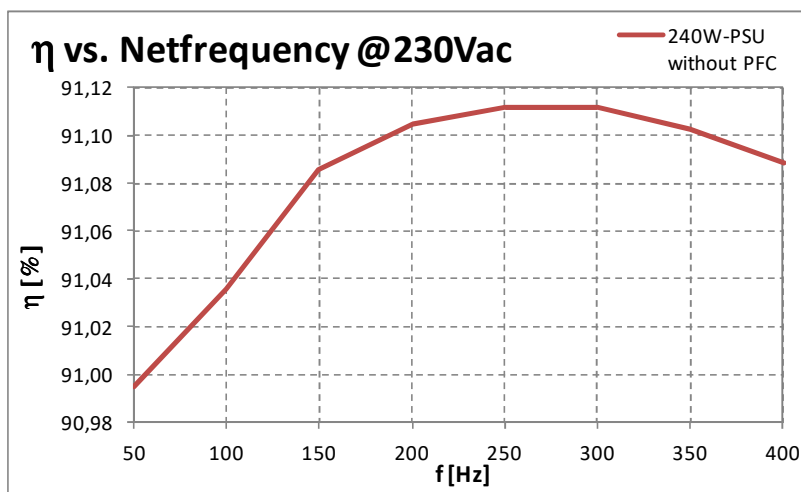


Fig. 4.18

Both units show up a small deviation of 0.1%. While the active PFC-controlled efficiency decreases slightly with an increasing line frequency (Fig), the non-PFC version seems to have an optimal efficiency at 250Hz.

Conclusion: In the usual range of 47-63Hz the effect of the mains frequency is negligible regarding efficiency.

4.2.4 3-Phase Balance

Since 3-Phase networks usually aren't loaded evenly by 1-Phase appliances mostly without PFC (see also section 4.2.1), deviations in line magnitude and phase angle occur frequently. This imbalance has considerable influence on the line input current of 3-Phase PSUs especially without active PFC. A simulation of a 240W PSU with passive PFC choke demonstrates phase failure L1 in case of voltage sag -10%. This leads to higher RMS currents in the input circuitry, increasing power losses and downgrading of efficiency.

Balanced network 400Vac, 50Hz, Pout=240W

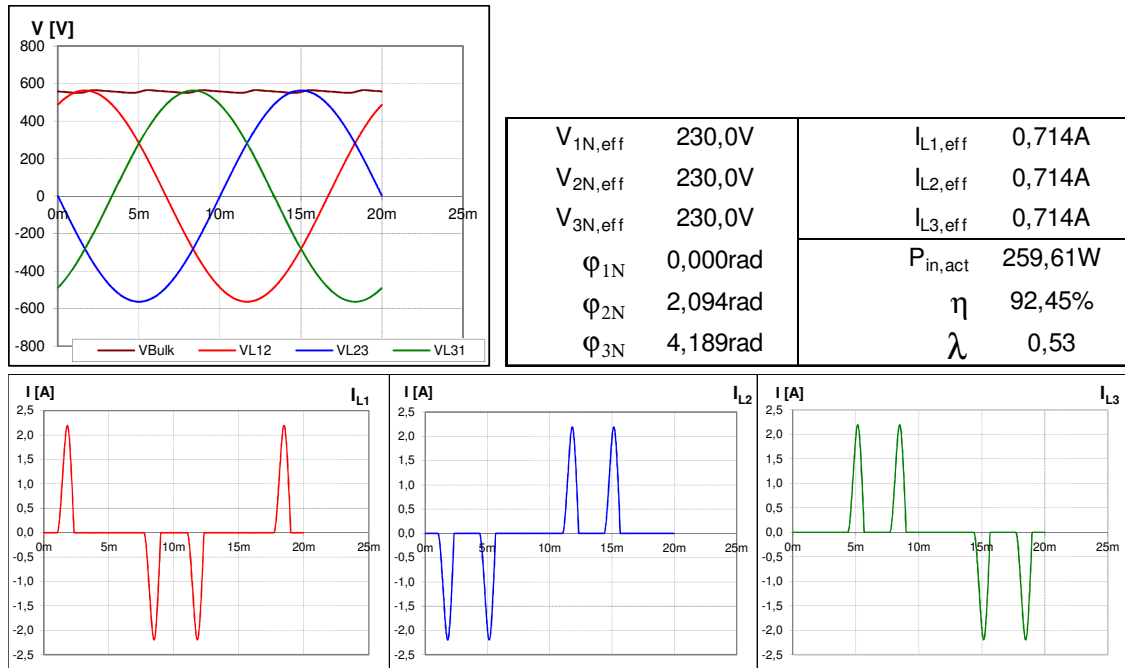


Fig. 4.19

Unbalanced network 400Vac, 50Hz, 1N -10%, Pout=240W

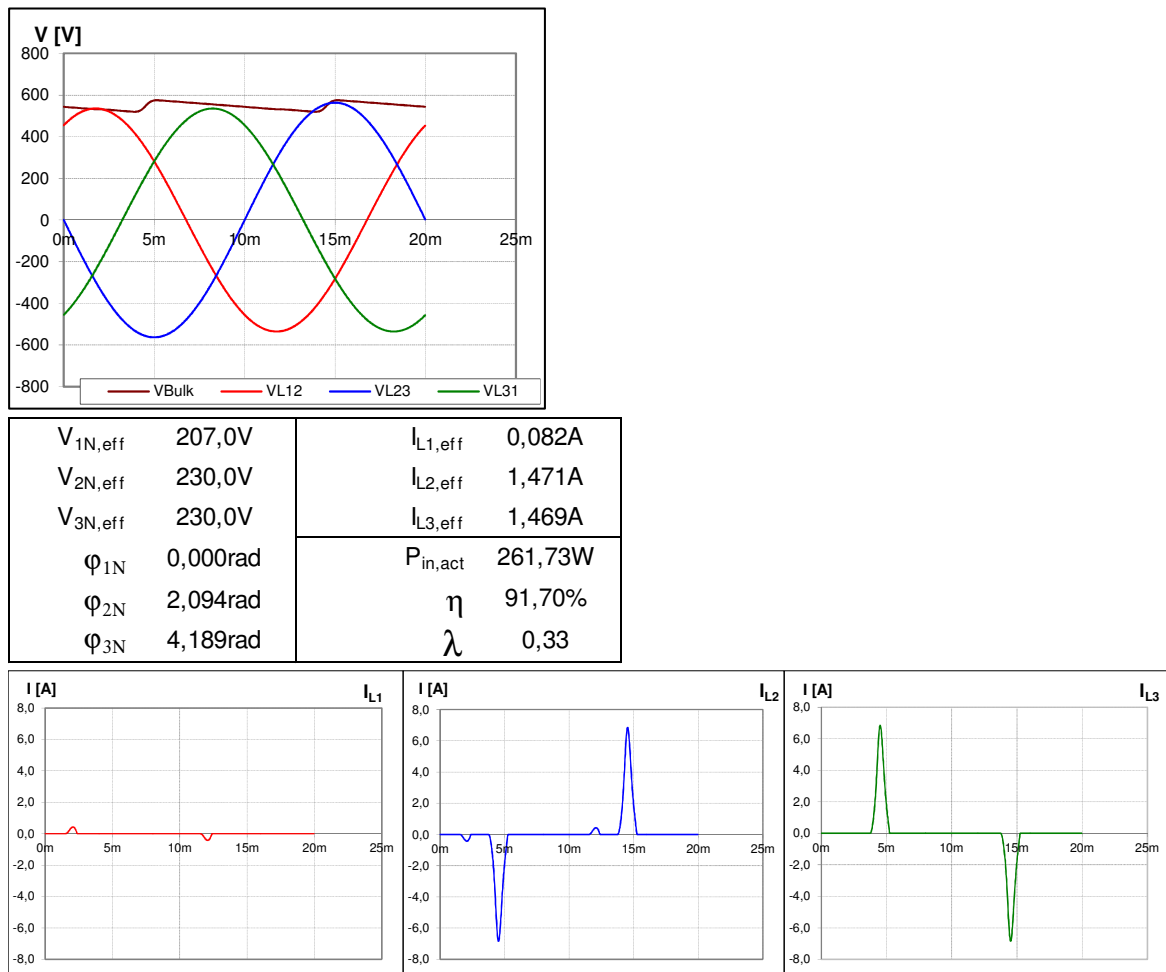


Fig. 4.20

Even more critical is the phase angle. Fig. 4.21 shows a deviation of -4° (-3.3% of symmetry) of V_{1N} leading to phase failure L3.

Unbalanced Network 400Vac, 50Hz, $\phi_{1N} -4^\circ$, $P_{out}=240W$

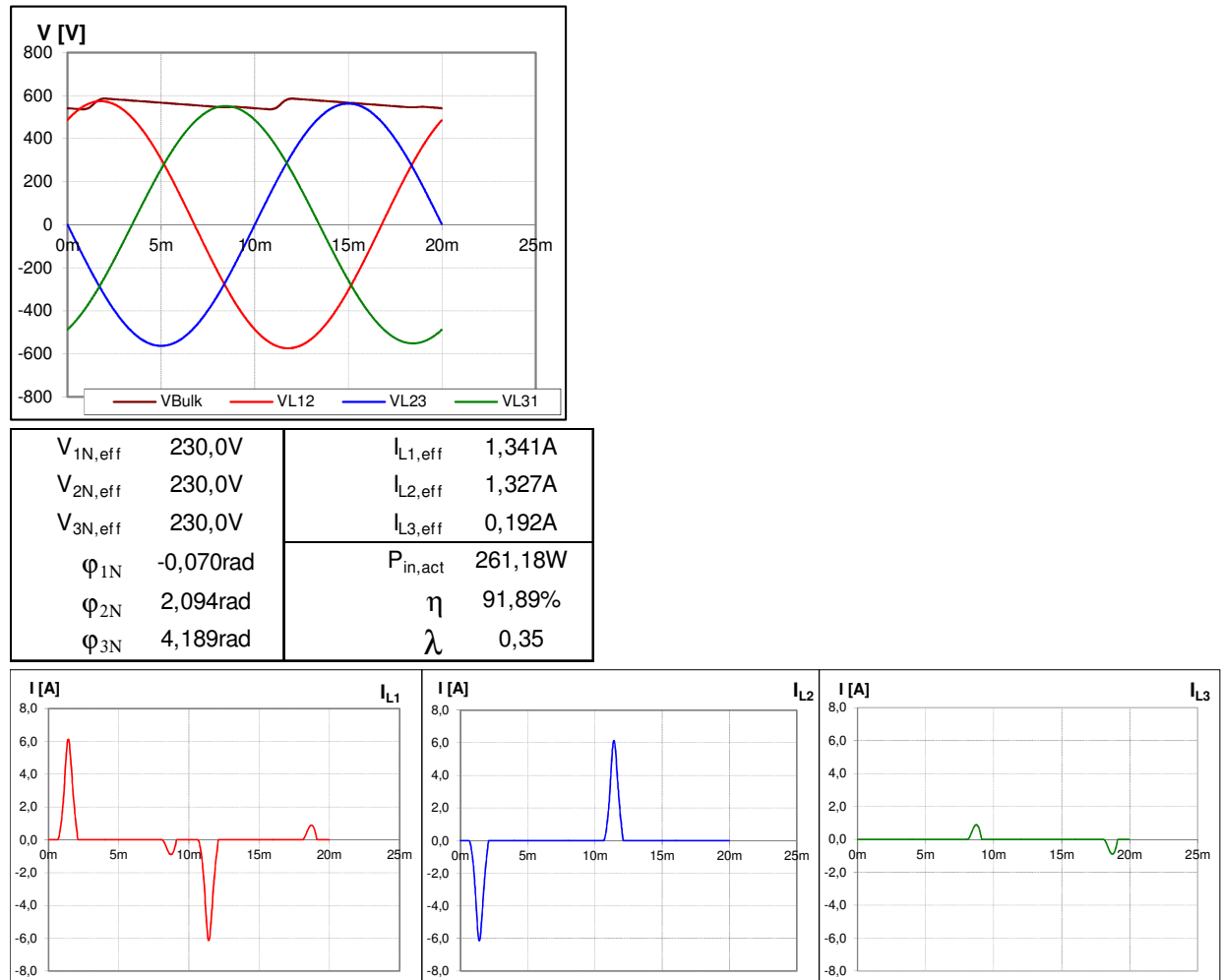


Fig. 4.21

Conclusion: Quite small deviations in balance of the power net results in phase failures raising rms currents in the input stage thus increasing power losses. The examples state a variation in input power of $+2.1W$ respectively -0.75% in efficiency. Since asymmetry basically downgrades efficiency, it should be of interest to use a 3-Phase AC power source to guarantee balanced current feeding of a 3-Phase PSU without active PFC. According to IEC61000-3-2 Annex A.2 "...the test voltage of each phase shall be maintained within $\pm 2.0\%$ and the angle between the fundamental voltage on each pair of phases of a three-phase source shall be $120^\circ \pm 1.5^\circ$ " to fulfill the harmonics regulation. These tolerances shall also be recommended for measuring efficiency in order to get reproducible results.

Units with active PFC principally don't have an issue since the current angle is independent of the voltage peak.

4.3 Simultaneous and sequential measurements

4.3.1 Propagation of Error

The total error of the general formula $m = f(x_1, x_2, x_3, \dots)$ can be derived by the total differential:

$$dm = \frac{\partial f}{\partial x_1} dx_1 + \frac{\partial f}{\partial x_2} dx_2 + \frac{\partial f}{\partial x_3} dx_3 + \dots$$

Formula 4.9

Whereas the absolute error can be described by:

$$\Delta m = \left| \frac{\partial f}{\partial x_1} \right| \Delta x_1 + \left| \frac{\partial f}{\partial x_2} \right| \Delta x_2 + \left| \frac{\partial f}{\partial x_3} \right| \Delta x_3 + \dots$$

Formula 4.10

Respectively the relative error (in %)

$$\varepsilon = \frac{\Delta m}{m} = \left| \frac{\partial f}{\partial x_1} \right| \cdot \frac{\Delta x_1}{x_1} + \left| \frac{\partial f}{\partial x_2} \right| \cdot \frac{\Delta x_2}{x_2} + \left| \frac{\partial f}{\partial x_3} \right| \cdot \frac{\Delta x_3}{x_3} + \dots$$

Formula 4.11

For non-correlated input parameters, the probable error (standard deviation) can be calculated by

$$\Delta m = \sqrt{\left(\frac{\partial f}{\partial x_1} \Delta x_1 \right)^2 + \left(\frac{\partial f}{\partial x_2} \Delta x_2 \right)^2 + \left(\frac{\partial f}{\partial x_3} \Delta x_3 \right)^2 + \dots}$$

Formula 4.12

Applying Formula 4.10 to the standard formula for DC-efficiency $\eta = \frac{V_{out} \cdot I_{out}}{V_{in} \cdot I_{in}}$, this results in an absolute error:

$$\Delta \eta = \frac{I_{out}}{V_{in} \cdot I_{in}} \Delta V_{out} + \frac{V_{out}}{V_{in} \cdot I_{in}} \Delta I_{out} - \frac{V_{out} \cdot I_{out}}{V_{in}^2 \cdot I_{in}} \Delta V_{in} - \frac{V_{out} \cdot I_{out}}{V_{in} \cdot I_{in}^2} \Delta I_{in}$$

Formula 4.13

The relative error in (%):

$$\varepsilon_\eta = \frac{\Delta \eta}{\eta} = \frac{\Delta V_{out}}{V_{out}} + \frac{\Delta I_{out}}{I_{out}} - \frac{\Delta V_{in}}{V_{in}} - \frac{\Delta I_{in}}{I_{in}} = \varepsilon_{V_{out}} + \varepsilon_{I_{out}} - \varepsilon_{V_{in}} - \varepsilon_{I_{in}}$$

Formula 4.14

The probable error is:

$$\Delta\eta = \eta \cdot \sqrt{\left(\frac{\Delta V_{out}}{V_{out}}\right)^2 + \left(\frac{\Delta I_{out}}{I_{out}}\right)^2 + \left(\frac{\Delta V_{in}}{V_{in}}\right)^2 + \left(\frac{\Delta I_{in}}{I_{in}}\right)^2}$$

Formula 4.15

Conclusion:

The total relative error of multipliers or divisors can be determined by adding or subtracting each factor's relative error.

4.3.2 Sequential Data logging

Sequential measurement of voltage and current is adequate for time invariant voltage and power, e.g. DC-DC conversion. Input/output voltage and current then can be characterized individually by average values over a long period and multiplied to get the input and output power. This facilitates an accurate acquisition of power.

By using the same voltmeter for V_{in} and V_{out} and the same current meter for I_{in} and I_{out} in a similar range a similar tolerance behavior can be assumed, thus according to Formula its systematic error is compensated.

Conclusion: The use of a data logging device is more accurate than individual voltage and current meters.

4.3.3 Simultaneous Sampling

Time variant real power in general can be specified over a period T by integrating the instant power

$$P = \frac{1}{T} \int_T V(t) \cdot I(t) dt$$

Formula 4.16

Simultaneous capturing of voltage and current is necessary for calculating the instantaneous power.

Sinusoidal voltage simplifies calculation of active power in the frequency domain according Formula 4.1 and

Formula 4.2 to:

$$P = S \cdot \lambda = V_{in,rms} \cdot I_{in,rms} \cdot \sqrt{1 - THD_R^2} \cdot \cos \varphi_1$$

Formula 4.17

Measuring apparent power and harmonic distortion indeed is time uncritical. But since the phase between the fundamental current and voltage is required, this procedure also demands a time-critical data acquisition of voltage and current.

When determining efficiency at time variant power, the ratio of instant power is not of interest, it is the energy balance over a period τ which is crucial:

$$\eta = \frac{W_{out}}{W_{in}} = \frac{\int_{\tau} P_{out}(t) dt}{\int_{\tau} P_{in}(t) dt} = \frac{\frac{1}{\tau} \int_{\tau} P_{out}(t) dt}{\frac{1}{\tau} \int_{\tau} P_{in}(t) dt} = \frac{\overline{P_{out}}}{\overline{P_{in}}}$$

Formula 4.18

Hence simultaneous sampling of instant input and output power isn't necessary while the average input and output power becomes stable over τ . In other words, efficiency can be specified by capturing the ingoing and outgoing energy within a period τ . τ should be selected to be long enough to get steady-state results.

Conclusion: Simultaneous sampling of voltage and current is essential for determining time variant instant power. Since the balance of incoming and outgoing energy within a period of time is relevant for efficiency, acquisition of averaged incoming and outgoing power can be done asynchronously, e.g. by a wattmeter. The average time selected should be long enough until stable steady-state is reached. If steady-state is not reachable due to unsteady load conditions, simultaneous measuring of the input and output power has to be done, e.g. by a power analyser.

4.4 Effect of ripple and noise

4.4.1 Common/Differential Mode Noise

Conducted differential mode noise in the form of voltage ripple at the input or output of the PSU is mainly caused by switched AC current from the PFC stage or converter causing voltage drops across the impedance of blocking capacitances. The noise spectrum consists of switching frequency harmonics up to several MHz. Mains rectifier diodes cause additional noise by their reverse recovery current. This spectrum consists of multiples of the mains frequency up to 500kHz visible as conducted EMI at the input.

Conducted common mode noise in the form of current spikes at the input or output is mainly caused by charge transfer due to switching voltage potential across parasitic capacitances to protective earth (PE) or to secondary like mounting capacitance of semiconductors or transformer winding capacitance. The so-caused noise spectrum extends over a wide range up to and beyond 30MHz.

Levels are regulated by EMI standards like CISPR 11/22, specifying Quasi Peak Class B up to 66dB μ V@150kHz/50Ohm. In the range of <150kHz other regulations like GL EMC1 allow levels up 80dB μ V@90kHz/50Ohm and a noise current of 200 μ A_{eff}. Common mode spikes are especially a possible irritator for the measurement equipment when evaluating different potentials (primary, secondary) using a datalogger or power analyzer, as parasitic capacitances between the

input channels could cause parasitic current into the measurement unit. Typical problems mainly occur while sensing shunt resistors, as the voltage is only in the range of some 100mV. These effects are known when testing PSUs which don't comply with EMI standards for conducted emissions. Blocking methods like ferrite clips on sense wires may stabilize the measurement.

Conclusion: For reliable efficiency measurements, standards for conducted EMI like CISPR 11/22 should be fulfilled. Otherwise avoid data logging of shunt voltages from primary and secondary.

4.4.2 Leakage current

PSUs have capacitances from line/primary to conductive earth parts like metal housings (Protection Class 1). These capacitances may be intentional, such as in an EMI filter to minimize common mode noise, or unintentional, like mounting capacitances of semiconductors, Transformers, PCB. The presence of AC potential to earth across these parts causes a current from Line to Earth, the so-called leakage current. This happens if the source has an impedance Z to Earth potential or is even connected like in a TN- or TT-Grid. Safety standards regulate the maximum, e.g. IEC950 <3.5mA (Class 1 with PE) or 0.25mA (Class 2 isolated).

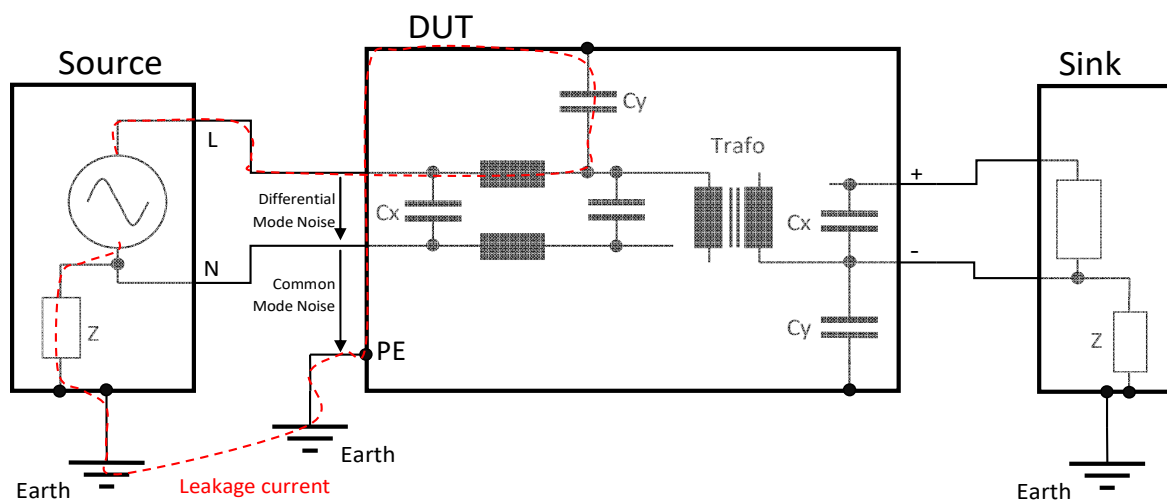


Fig. 4.22

Fig. 4.22 shows leakage current only occurring in line L and returning via Earth, not visible in the neutral N. By measuring low-powered DUTs, for e.g. standby losses with input currents of a few mA, the leakage current can't be neglected. It has to be determined where to measure the input current. Line delivers the total current. The problem: It is at high potential, therefore dangerous for the measurement equipment. Adequately isolated wires and current probes should be used. Measuring the current in the neutral, N, on the other hand is safe because of the low potential, especially preferable for shunting, but leakage current then is not visible! Another common way is to connect the metal housing to N or 'N'-pole, at least the leakage current to the housing is then visible. The problem then: Other Impedances from housing to Earth like fixture and load on secondary cause further bypass to earth, making the efficiency measurement unrepeatable.

Conclusion: The only reliable way to measure the input power is to connect the PSU's Earth (Class 1) and measure the current in L to also capture leakage current.

4.5 Environment

4.5.1 Thermal influence

Fig. 4.23 shows that once steady state is reached after approx. 30min, the efficiency of an old-fashioned DIN Rail PSU with free convective heat flow is indeed lower but almost independent of ambient temperature in the mid-range. Parts with positive and negative temperature coefficient t_c to keep the power losses in balance. At high T_{amb} an increase in power losses can be noticed due to dominating ferrite losses.

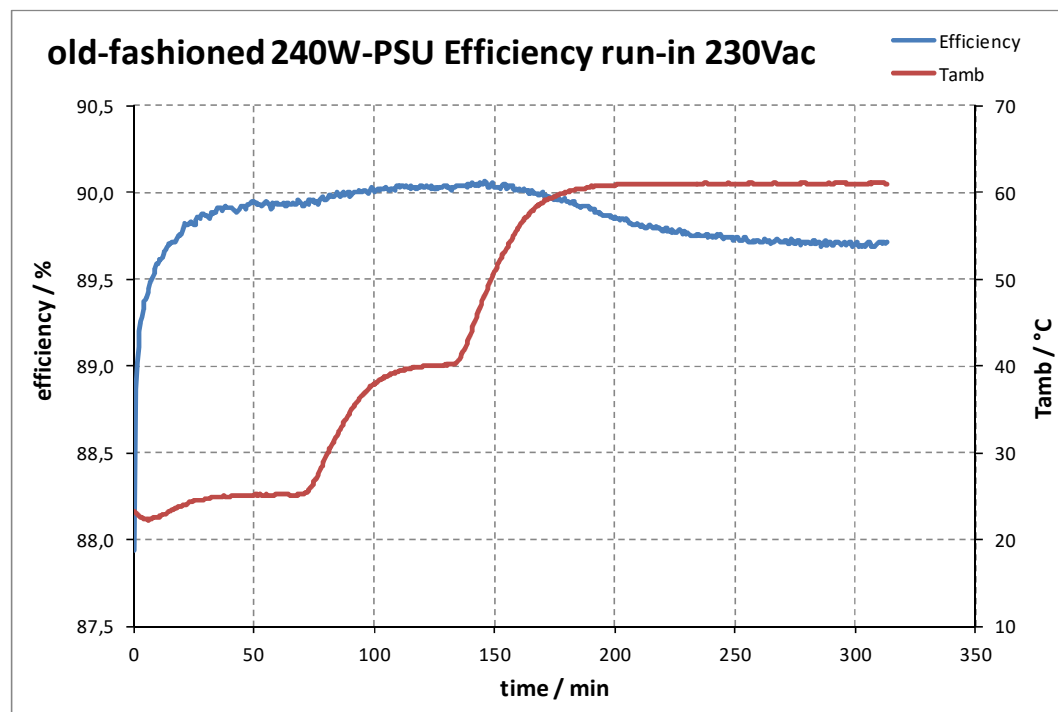


Fig. 4.23

In order to minimize losses nowadays, PSUs utilize more semiconductors which happen to have positive t_c in resistance like MOSFETs and SiC-Diodes instead of conventional silicon/schottky rectifier diodes with a negative t_c in voltage drop. Application examples are synchronous rectification secondary and bridgeless PFC primary. Switched inrush solutions instead of a simple but lossy inrush-NTC also increase η significantly but enforce positive temperature drift of the total power losses. PSUs typically still include parts with significant negative t_c like ferrites (up to 100°C) and electrolytic capacitors.

Fig. 4.24 and Fig. 4.25 demonstrate a typical efficiency response of a highly efficient 240W PSU without inrush NTC, convective heat flow by stepping T_{amb} 25°C / 40°C / 60°C

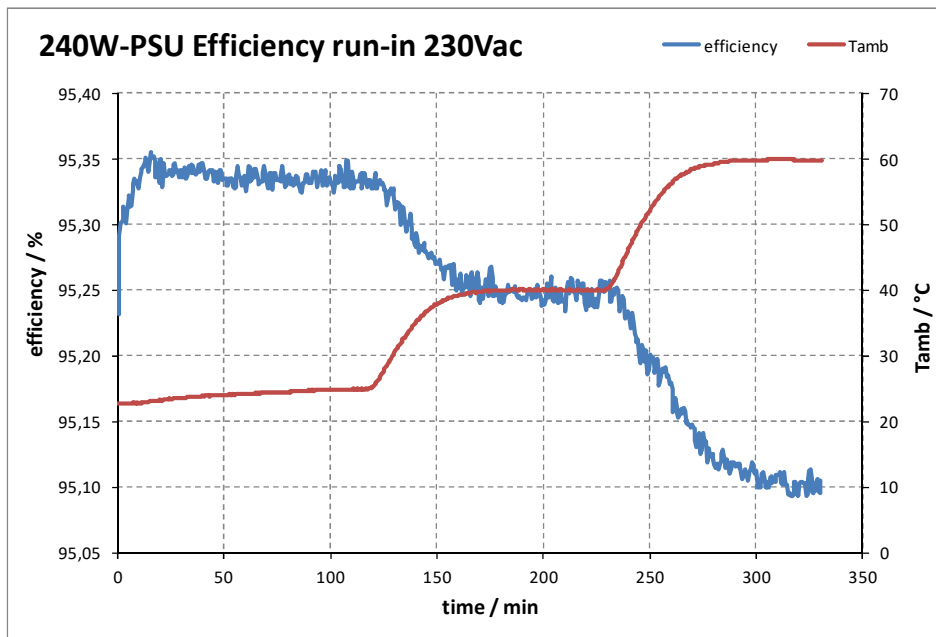


Fig. 4.24

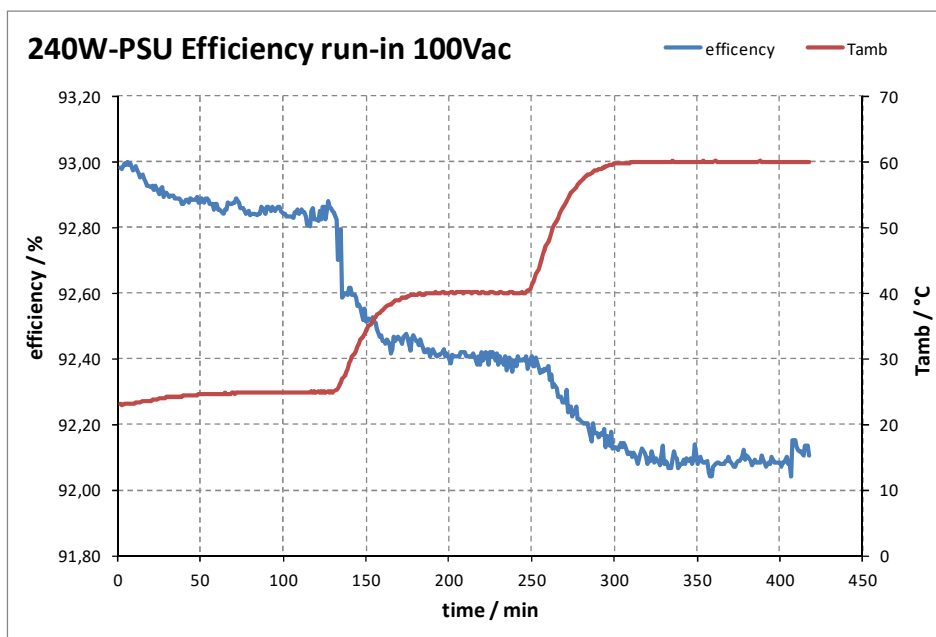


Fig. 4.25

It can be noticed from Fig. , at 230Vac the efficiency looks quite balanced, an acceptable deviation of 0.2% in the range 25-60°C, whereas at 100Vac (Fig. 4.25) there is a significant decrement of 0.8% in η . Steady state is finally reached after 1 hour.

Obviously, power losses caused by resistive parts with positive t_c dominate more at 100Vac than at 230Vac. In the worst-case, thermal steady-state can't be established, the unit fails!

Conclusion: Factors influencing the thermal state, like ambient temperature, mounting situation and atmospheric pressure (altitude) are of interest while measuring efficiency. Thermal steady-state basically has to be reached (settling time of at least 30min). Tamb has to be documented in the protocol, usually 25°C. Even the sequence of measured operating points is of interest, as the thermal steady-state could deviate.

4.5.2 Altitude

The barometric pressure correlates with the altitude h by the barometric formula for standard atmosphere

$$p(h) = 1013,25hPa \cdot \left(1 - \frac{0,0065 \cdot h}{288,15m}\right)^{5,255}$$

Formula 4.19

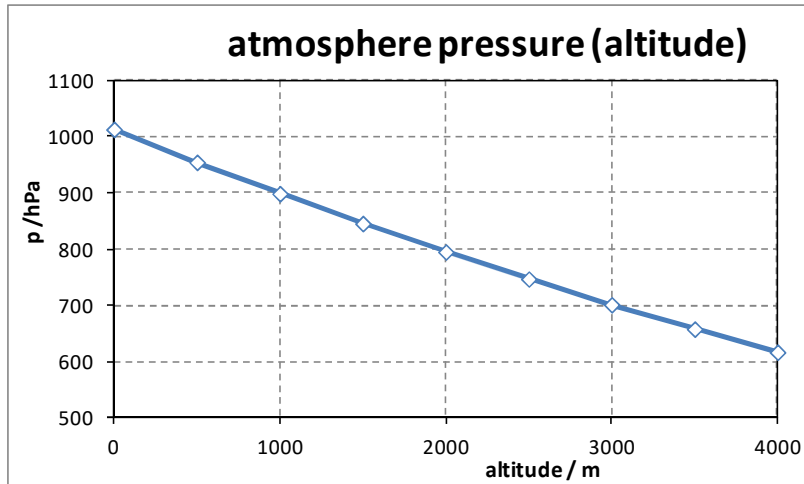


Fig. 4.26

Roughly speaking, atmosphere pressure reduces linearly by 100hPa/1000m.

Fig. 4.27 (Source: H. Saidi, R.H. Abardeh, WEC 2010, London) shows a practical thermal investigation with a heated copper cylinder placed horizontally in a pressure chamber. Air pressure and copper temperature influence significantly convective heat transportation.

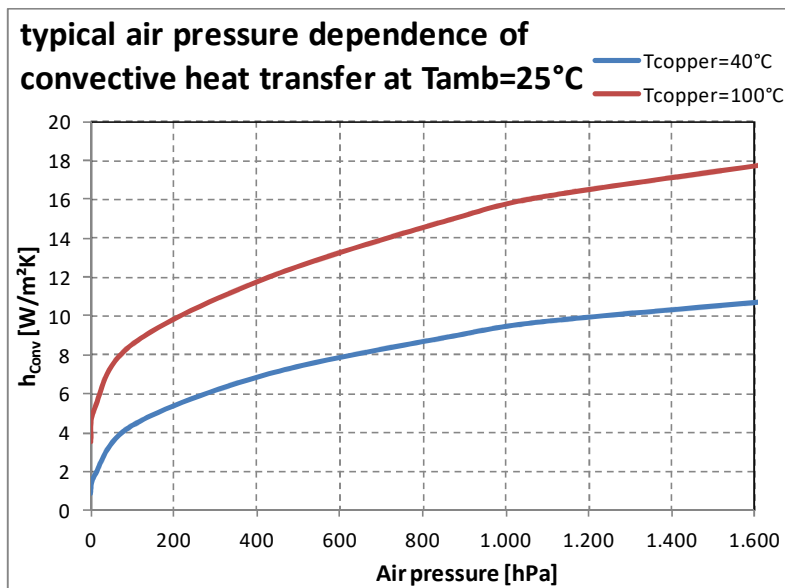


Fig. 4.27 (Source: H. Saidi, R.H. Abardeh, WEC 2010, London)

A heat transfer reduction of -8% @800hPa (2000m altitude) compared to sea level can be stated, respectively -12% @700hPa (3000m).

A thermal simulation of a 240W PSU with free convective air flow at $T_{amb} = 55^{\circ}\text{C}$ shows an even higher thermal rise:

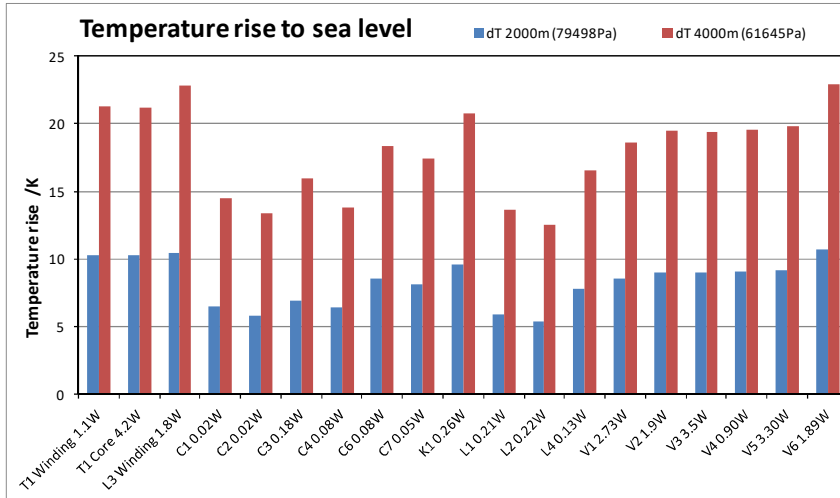


Fig. 4.28 Simulated temperature rise caused by change of air pressure

Due to less convective heat transportation, a temperature rise of almost 10K @2000m and even 20K @4000m can be noticed increasing power losses in parts with positive temperature coefficient like MOSFETs and copper. Accordingly, the efficiency at $T_{amb} / 2000\text{m}$ in this 240W PSU-example is comparable to sea level with an increased $T_{amb} +10\text{K}$ (see section 4.5.1), which means according to Fig. 4.25, 0.4% less efficiency at 100Vac!

Conclusion: Since air pressure influences η the measurement protocol should contain a declaration of the test altitude.

4.5.3 Humidity

Humidity could lead to condensation on parts impacting the electrical performance of a PSU by creepage current. This may influence the functionality and cause additional losses.

An influence of humidity on convective heat transportation changing the thermal profile is not detected either by thermal simulation or by any variation in efficiency measured with a 240W DIN Rail PSU with free convective air flow @ $V_{in}=230\text{Vac}$, $P_{out}=240\text{W}$:

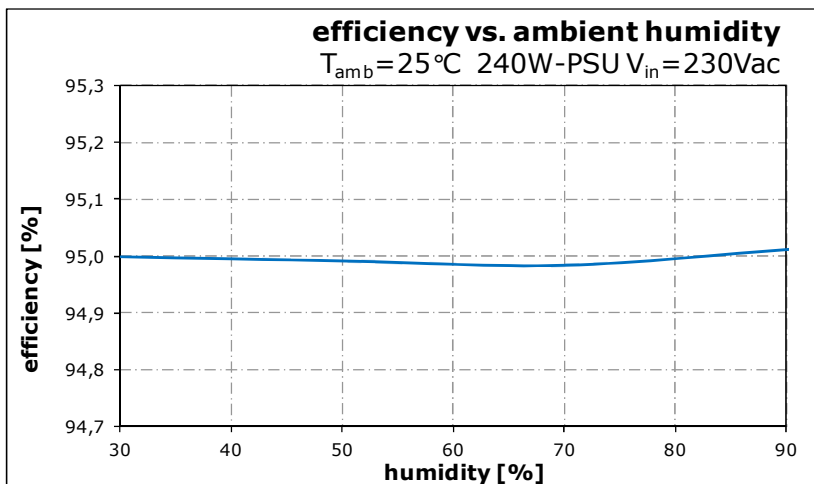


Fig. 4.29

Conclusion: Humidity doesn't seem to affect efficiency measurement, hence can be neglected unless condensation is involved.

4.6 Measurement setup

4.6.1 Wiring / contact resistance

Wiring and contact resistance cause additional losses therefore have to be excluded systematically from the power measurement. This is mainly accomplished by sensing input and output voltage directly at the terminals of the DUT.

In extreme situations, contact losses at the terminals can raise the thermal profile in the PSU, which could lead to higher power losses inside. Adequate wire diameter should be selected.

In testing AC PSUs without PFC, all sorts of impedances in the input circuitry as described in sections 4.2.1 and 4.2.2 reduce the input rms current and thus lessen the power losses inside the DUT resulting in an increase in efficiency. The total wiring resistance shouldn't be exceeded as shown in Formula 4.8.

4.6.2 Sensing Input/output

For determining input and output power, incoming and outgoing voltage and current has to be detected correctly. For measuring power >10W, voltage should be sensed directly at the terminals of the DUT, otherwise voltage error caused by R_i , wiring and contact resistance falsifies power calculation.

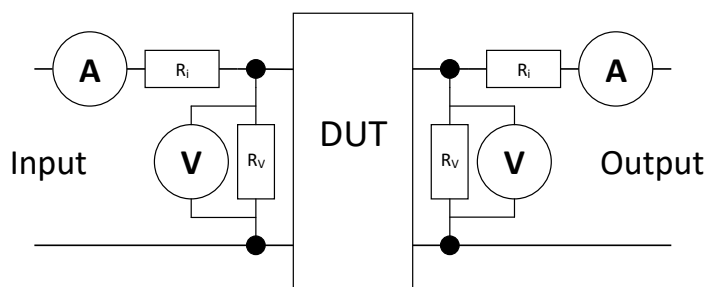


Fig. 4.30

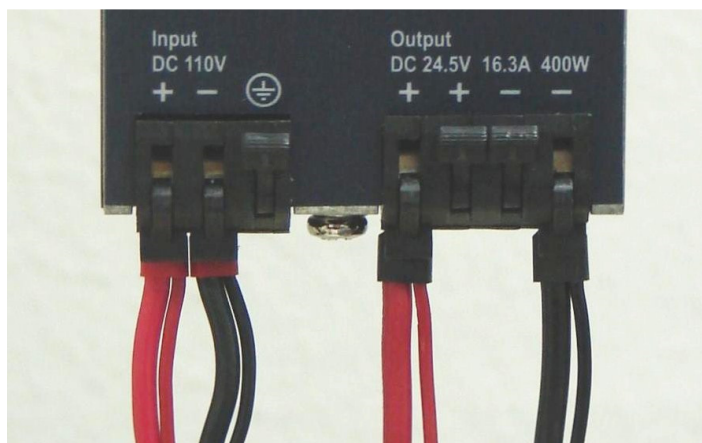


Fig. 4.31

If more than 2 terminals are available, separating power and sense terminals may be of advantage to exclude contact resistance at the terminals. Built-in voltmeters in sources and sinks may only be used with additional sense wires.

In the case of determining power <10W, e.g. standby power losses, current error caused by voltmeter impedance R_v (typ. >1M Ω) has to be taken in account, whereas voltage error can be neglected. It is therefore recommended to exclude the error current by a vice versa connection.

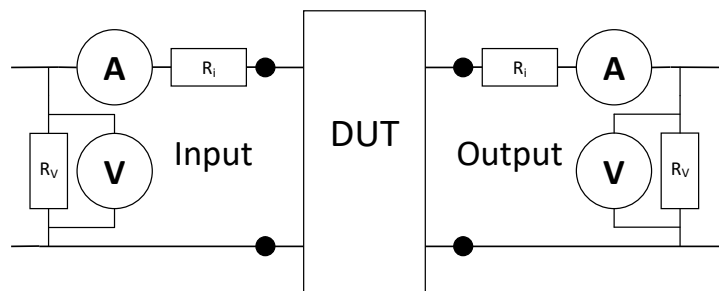


Fig. 4.32

Current measurement at the input and output should be done either with a shunt or current probe. Built in current displays in source and sink are usually too inaccurate. The resistance of a shunt should be selected suitable to the scaling of the voltmeter and not too high in order not to influence the power factor especially of devices with no PFC at the input (see Formula).

4.6.3 Load (type and sequence)

As efficiency of a PSU depends on its operating point, reproducible load conditions are essential. Basically, any sort of load on a DC Output is suitable for measuring efficiency, as long as it can guarantee stable output power.

I-Load stabilizes the load current, hence is appropriate for PSUs with voltage-controlled outputs. Outstanding advantage: Load wiring cannot influence the current, therefore a preferable method. Inaccuracy in the regulated output voltage influences the output power linearly.

Z-Load stabilizes the load voltage. This type is preferred for current-controlled outputs like LED drivers. Voltage drop in the wiring has to be considered, thus voltage sensing at the output terminals is necessary to guarantee stable load conditions. Inaccuracy in the regulated output current influences the output power measurement linearly.

R-Load defines a linear V-I behavior. Neither voltage nor current is defined. Easy to realize by ohmic resistances, can be used for voltage- and current-controlled output, but does not guarantee a stable output power, since the resistance of the resistive load varies according to its temperature rise and the output wiring also affects results. Inaccuracy in the regulated output voltage or output current influences the output power to the power of 2, therefore not recommended for efficiency measurement.

4.6.4 Test Sequence

It is known that heating up of power supplies to a certain T_{amb} could lead into a different thermal steady-state than cooling down to T_{amb} . Usually the cooling down response time lasts longer than heating up. As the thermal operating point has consequences for the power losses, the test procedure for different operating points should be stated basically according to an overall increasing thermal profile:

- T_{amb} increment starting at lowest temperature
- V_{in} decrement starting at highest input voltage
- P_{out} increment starting at smallest output power

4.6.5 Mounting position, enclosure

The mounting arrangement affects in particular PSUs without forced airflow. As free convective heat flow is influenced by different mounting positions, the thermal profile changes, thus losses and efficiency change. A test series with an actual DIN Rail 240W PSU with free convective air flow shows however only slight deviation in efficiency of <0.1%:

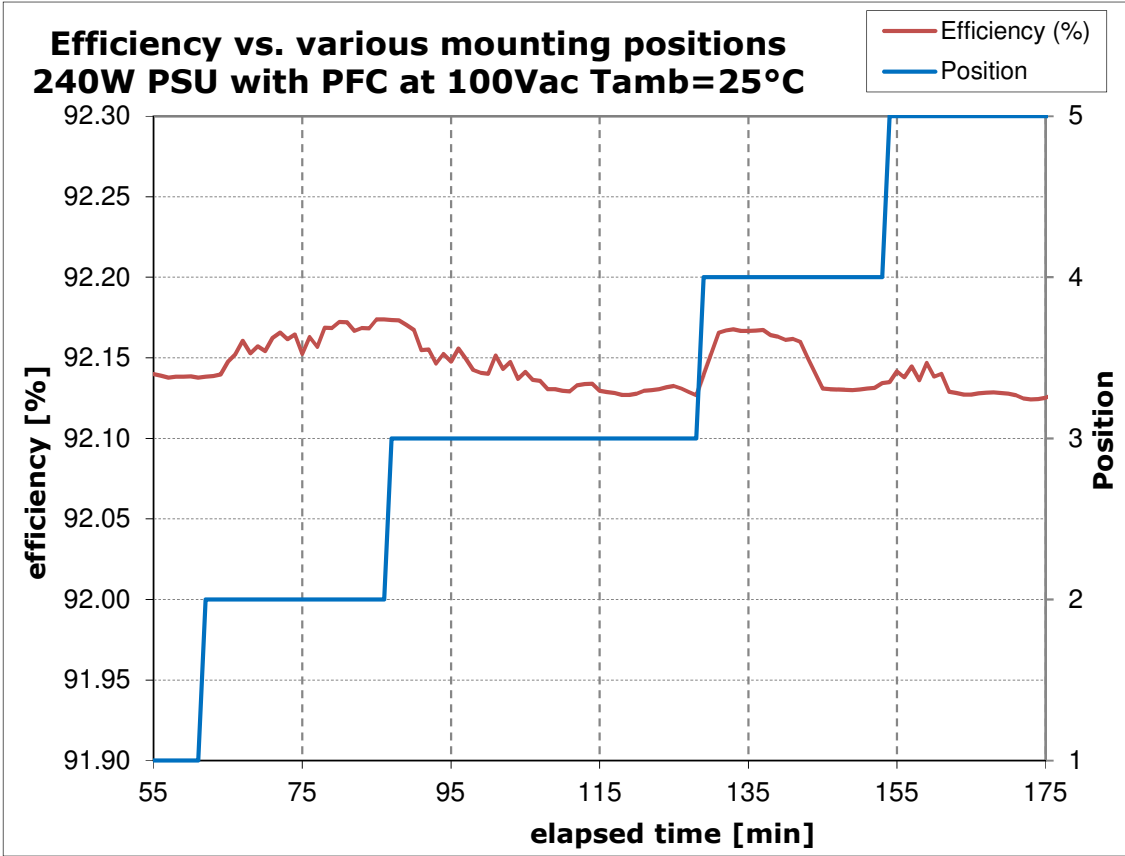


Fig. 4.33

Pos	
1	DIN Rail regular
2	DIN Rail upside-down
3	Table top
4	Table top horizontal CW
5	Table top horizontal CCW

Nevertheless, the typical application environment should be applied, e.g. DIN-Rail, 19"-rack, customer specific cabinet. Unless otherwise specified, forced airflow surrounding the DUT has to be avoided by e.g.:

- Closed door and window
- Switched-off fan in climate cabinet
- Surrounding test box

5 Proper instrumentation

5.1 Multimeter

The error Δm of a measured value m in general can be described as follows, wherein p_m is the percentage error of the reading value m and p_r the percentage error of the full-scale value m_r
 Absolute value:

$$\Delta m = \text{readingErr} \text{ or } + \text{rangeError} = p_m \cdot m + p_r \cdot m_r \tag{Formula 5.1}$$

Relative value (in %):

$$\frac{\Delta m}{m} = p_m + p_r \cdot \frac{m_r}{m} \tag{Formula 5.2}$$

The range error of a digital multimeter is usually specified by its resolution in counts and its number of least significant digits n_{lsd} :

Relative error (in %):

$$\frac{\Delta m}{m} = p_m + \frac{n_{lsd}}{\text{counts}} \cdot \frac{m_r}{m} \tag{Formula 5.3}$$

Tricky to handle is the resolution. It depends on how the full-scale value fits in the number of decimal digits in the display ("display count"). E.g. a 3½-digit voltmeter (reading error 0.5%) with a voltage range up to 6V fits in a display with only 3 decimal digits, hence the resolution is 600 counts (0-5.99V). Whereas a voltage range up to 200mV fits in 4 decimal digits, the resolution is 2000 counts (0-199.9mV). This link to an Excel-Tool gives an idea how ranging impacts the error of measured values: <http://www.epsma.org/efficiency.html>

A right choice of the measuring range is obviously essential for improving accuracy.

The meaning of the fractional for the most significant decimal digit is not explicit, commonly $\frac{1}{2} = "1"$, $\frac{3}{4} = "1,2,3"$, $\frac{4}{5} = "1,2,3,4"$.

5.2 Datalogger

A datalogger is a useful multiplex meter, which can replace several individual multimeters.

Thereby systematic error can be compensated according to section 4.3.2 when measuring DC-DC efficiency hence improving accuracy significantly.

Dataloggers like the Agilent 34970 are equipped with a superior, precise 6½-digit measurement unit differing 0.004% of reading and 0.0007% of range (1V-range) and a multiplexing unit. Since relays are used for switching the channels, aging of the contacts is an issue which impacts the accuracy allowing only temporary tolerance specification up to 1 year. Consequently, dataloggers have to be calibrated frequently.

As mentioned in section 4.4.1 common-mode disturbances have to be considered when testing isolated DC-DC converters, because these disturbances could cause errors in measurements.

5.3 Wattmeter

A wattmeter is quite affordable but rather inaccurate (around 1%) in capturing time variant electric real power over a period of time, whereas for measuring DC power the use of a multimeter for measuring voltage and current (with shunt) is more accurate. So, if possible the use of a wattmeter should be avoided.

Specific devices are available to calculate the electric energy for metering purposes, which are based on integrated circuits, available from various companies. The critical task in getting real power is to realize a real time 4-quadrant multiplication $V_{load}(t) \cdot I_{load}(t)$ with a succeeding averaging. Since power is mainly measured at mains sine voltage, typically 50-60Hz, a rather narrow frequency spectrum up to 10kHz is of interest, hence several methods are adequate: Traditional electrodynamic meters. The multiplication is done by I_{load} inducing B in a fixed current coil and a moving coil fed by I_{sense} proportional to U_{load} (potential coil). The deflection of the pointer is caused by the Lorentz force:

$$\vec{F}_L = I_{sense} \cdot (\vec{B}(I_{load}) \times \vec{l})$$

with its inertia realizing the average power. This method is suitable also for highly distorted currents and voltages (Crest-factor < 10). However, since range error is in the percentage range due to mechanics, and reading error is a matter of individual reading skills hence not specified, analog measuring equipment is basically not suitable for gaining accurate and reproducible results.

Hall effect multipliers are also based on the Lorentz force. By feeding the coil with I_{load} and the transducer with I_{sense} a Hall-voltage (representing the instant power) is generated at the sensor

$$v_H(t) = R_H \cdot B(t) \cdot I_{sense}(t)$$

with $B \sim I_{load}$, $I_{sense} \sim V_{load}$, R_H = Hall constant.

The hall voltage finally has to be low pass filtered to obtain an average real power.

Hall sensors are quite accurate (tolerance < 0.2%) and easy to use, since B can be detected contactlessly from the wire with a current clamp. Peak current capability due to saturation has to be considered.

Time division modulation (TDM): PWM representing the signal amplitude~current and the duty cycle~voltage. The pulsed area correlates with the instant power, a following low pass filter reveals the real power. This method is rather slow, limited bandwidth only up to 20kHz, but very accurate (<0.05%) and works 4-quadrant.

Log-antilog-amplifier: The principle is to produce the logarithm of current and voltage so that the product (division) can easily be realized in the logarithmic domain by a simple addition (subtraction). The sum finally has to be exposed and low pass filtered to get real power. Natural log and antilog conversion can be done by using the logarithmic V(I) characteristics of a PN-diode or exponential IC(V_{be}) behavior of bipolar transistors. In order to match both conversions for accomplishing acceptable error and temp drift, this method can only be realized by monolithic chip integration, e.g. FAI RC4200 Single quadrant/0.2%/ 4MHz, AD 538 2-quadrant/0.5%/400kHz.

Translinear multiplier (Gilbert Cell): 4-Quadrant multiplier based on the fact that the transconductance (gain) of a bipolar transistor is linearly dependent on the IC

$$\frac{dI_C}{dV_{BE}} = \frac{q}{kT} \cdot I_C$$

Formula 5.4

By varying IC(t) the gain can be modulated. Acceptable tolerance and temperature stability is only realizable with monolithic chip integration, e.g. MC1495 2%/10MHz, AD633 1MHz/1%. This method is commonly used, but not as accurate as previous methods.

Digital solution with digital signal processing (DSP). By digitalizing current und voltage, all sorts of mathematical operations can be processed digitally like active/ reactive/ apparent power, rms, harmonics and power factor. This is a comprehensive way to evaluate current and voltage, mostly used in power analysers. When analysing a spectrum up to 10kHz, a sampling frequency >20kHz is necessary according to the Nyquist criteria.

For any AC or transient condition, the average Power is calculated by

$$\bar{P} = \frac{1}{t_2 - t_1} \cdot \int v(t) \cdot i(t) dt$$

Formula 5.5

By using an individual wattmeter at input and/or output, the measured input and output power doesn't correlate. A wattmeter can only be reliably used, if we have a steady energy flow over a period of time, which has to be guaranteed by stable source and load conditions (see also section 4.3.3). If power varies at the output e.g. during load stepping, the measurement gets much more complicated, since incoming and outgoing energy has to be captured in the same period of time. A more sophisticated power analyser has to be used instead, which measures input and output power simultaneously.

A simple wattmeter may not have any voltage sense input, hence wiring to the PSU has to be taken into account!

Conclusion: The use of wattmeters should be avoided due to inaccuracy and uncorrelated capturing of input and output power. Alternatively, for DC power, a voltmeter and current meter and for AC-power a power analyser should be used.

5.4 Power Analyser

Most accurate and reliable efficiency results are delivered by power analysers. The selection criteria are first of all accuracy, power, but also the number of available channels, to be able to make measurement from single phase to 3-phase products, but also within the power supplies. Simultaneous capturing of input and output power gives the possibility to measure the efficiency at time variant load conditions.

For accurate efficiency measurement in the range of getting close to the ideal 100% the theoretical maximum achievable accuracy is key. Best in class equipment like the Yokogawa WT3000 offer a base accuracy @50Hz for voltage and current measurements of 0.01% of reading, +0.03% of range and for power measurements 0.02% of reading + 0.04% of range.

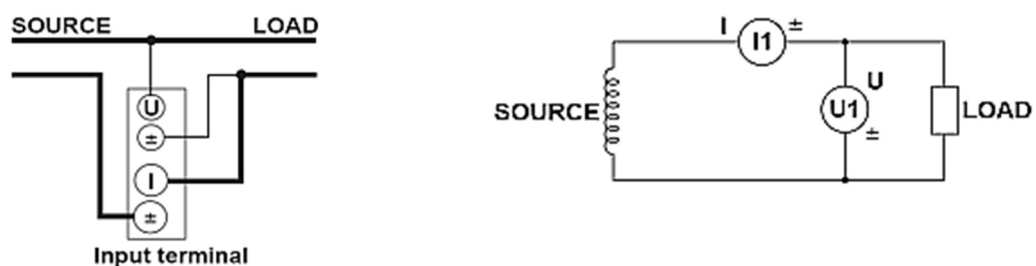


Fig 5.1: Typical connection scheme to the power analyzer's built-in voltage and current inputs (reference Yokogawa WT3000 manual)

Usually a multi-channel power analyser comprises n voltage channels and m current channels. For voltage measurement and current measurements up to a certain level, the built-in channels can be used. For higher current levels an extension sensor, either a shunt or clamp type sensor, can be used.

5.5 Shunting

Measuring current with an additional shunt resistor is usual for capturing Power at DC Voltage. Under these circumstances electric power can be determined using independently a current and voltmeter according to the equation:

$$P = V_{DC} \cdot I_{arith}$$

which is more accurate than a wattmeter. Since a multimeter's built-in current meter is too inaccurate in many cases and not capable of measuring high current >10A, sensing current with an additional shunt resistor is of advantage. Arithmetic averaging (not RMS!) of time variant current can be realized with the voltmeter, or by an additional current filter, which shouldn't cause significant additional losses.

How to select a suitable resistance value? On the one hand, the sense voltage shouldn't be too small as the range error of the voltmeter becomes too great (see section 5.1). On the other hand, too high resistance could overheat and cause excessive voltage drop influencing the operating point of a PSU. Power rating and thermal conditions of R_{shunt} regarding temperature drift also affect the measurement accuracy. Usually values range between 1-100mOhm. Resistance tolerance should be $\leq 0.1\%$, temperature drift $\leq 10\text{ppm/K}$. Only a precision

resistance with voltage sense terminals is suitable. Cooling the package with a heat sink reduces temperature rise and minimizes drift.

The impact of the shunt resistor's tolerance and range error to total efficiency accuracy is demonstrated at this link: <http://www.epsma.org/efficiency.html>

For correct sensing of the voltage see section 4.6.2.

Where to insert the shunt? For safety reasons it's recommended in the wire with the lowest potential to earth, usually (-) or N. But keep leakage current in mind (see section 4.4.2).

5.6 Power Source / Sink

Electronic power sources and power sinks are vital for measuring reproducible results, since they guarantee reproducible conditions. Especially when testing PSUs without PFC, the source impedance has a significant effect on the power losses inside the DUT (see section 4.2). Also, beware of current clipping since peak current may possibly be up to 5x RMS.

As mentioned in section 4.2.4 the balance in magnitude and phase of the 3-Phase feeding source is crucial to achieve a reasonable efficiency when testing 3-Phase PSUs without active PFC. A minor error of 10% in amplitude or 4% in phase could lead to 2-phase mode.

Stable load conditions have to be achieved when measuring input and output power uncorrelated e.g. with a wattmeter.

It is usually convenient to adjust to appropriate voltage and current levels using the built-in voltage and current control and the respective displays, or by automatic remote control. Since the built-in measurement devices are usually of lower precision with typical accuracy figures ranging up to 0.5... 1% each, accurate efficiency measurements are not feasible using that type of instrument, because the error can easily end up in a range of 2...4%, so as an exception and only if a 4-wire measurement for the voltage is in place, very coarse measurements are tolerable.

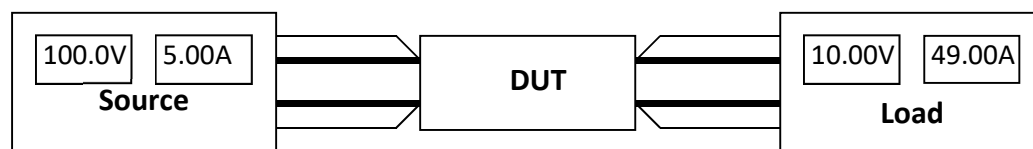


Fig 5.2: Measurement setup using voltage sensing at the DUT input and output for coarse efficiency measurements

6 Calorimetric methods

6.1 Background

In order to obtain the efficiency of a device, the usual way is to measure the incoming and outgoing power and to calculate the efficiency by formula 4.18.

Whenever the efficiency is approaching very high numbers close to 100%, the losses get very dependent on the accuracy of the measurement of the input and output power, which leads to large errors both in calculation of the losses as well as the efficiency figure itself (see also chapter 3).

In this case, the calorimetric method offers a very large benefit, since the losses are being measured directly and the power can be measured on either side of the DUT. In case of a power conversion device that converts AC to DC power, measuring the DC power is very easy and can be achieved with simple instruments, a volt-meter and an ampere-meter.

6.2 Methodology

For measuring the losses directly, the calorimetric method works as follows:

- Encapsulate the device under test (DUT)
- Establish a single heat transport
- Prevent any secondary heat transport
- Measure the transported heat

In a status of thermal equilibrium, the transported heat is equal to the power losses

The general heat, which is equivalent to the power loss in the DUT can be expressed as:

$$\partial Q = C_T(V) \cdot \partial V + C_V(T) \cdot \partial T$$

Formula 6.1

where the differential in heat ∂Q is dependent on volume differential and temperature differential. In order to simplify the measurement a constant volume calorimetric is being chosen, where the dependability on volume $\partial V \rightarrow 0$ can be neglected.

The remaining equation of Formula 6.1 is therefore:

$$\partial Q = C_V(T) \cdot \partial T$$

Formula 6.2

The specific heat at constant volume is dependent on the material as well as the temperature, whereas the second one can be kept at a minimum, when the appropriate material for the temperature transfer is chosen. As an additional means the temperature can be kept at a minimum by keeping the difference in temperature low and therefore the specific heat becomes a simple constant:

$$\partial Q = C_{mat} \cdot \partial T$$

Formula 6.3

which is only a material-dependent figure.

With the measured power losses of the calorimetric method, the calculated efficiency is:

$$\eta_{calorimetric} = \frac{P_{out}}{P_{in}} = \frac{(I_{out} \cdot U_{out})}{(I_{out} \cdot U_{out} + P_{calorimetric})}$$

Formula 6.4

Specific heat of transformer oil:	1860 J/(kg*K)
Purified water:	4187 J/(kg*K)

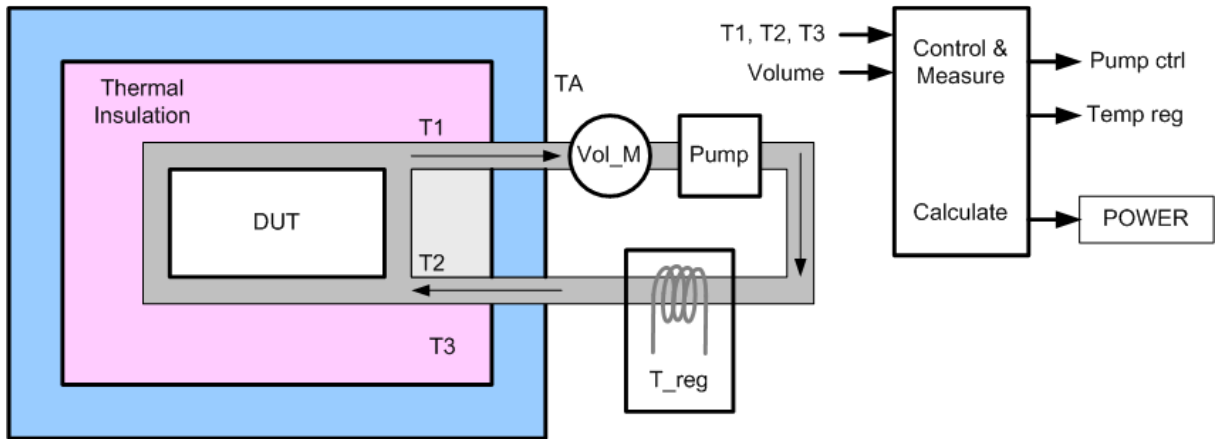


Fig 6.1: Basic scheme of a calorimetric measurement setup

6.3 Accuracy

Unlike the direct measurement of the real electric power at the input and output, where the losses are a small difference of relative large measures, the calorimetric method measures the losses directly by simply multiplying a volume and a temperature difference with a material constant.

Hence the accuracy of the measurement result is the product of the error:

$$\mathcal{E}_{temp}$$

in the temperature measurement with the error:

$$\mathcal{E}_{vol}$$

in the measurement of the volume of the heat transfer substance, mostly a liquid like purified water or transformer oil:

$$\mathcal{E}_{tot} = (1 + \mathcal{E}_{temp}) \cdot (1 + \mathcal{E}_{vol}) - 1$$

Formula 6.5

Hence:

$$\mathcal{E}(\eta_{calorimetric}) = \frac{1 + \mathcal{E}(P_{out})}{1 + \mathcal{E}(P_{out}) + \mathcal{E}(P_{calorimetric})}$$

Formula 6.6

If we now compare with the traditional method of the input and output power measurement, the power losses are:

$$P_{loss} = P_{in} - P_{out}$$

Formula 6.7

And the maximal error:

$$\varepsilon(P_{loss}) = \varepsilon(P_{in}) + \varepsilon(P_{out})$$

Formula 6.8

which is dependent on the ratio of the power loss and the output power.

Measurements at high efficiency levels clearly visualize the benefit of the calorimetric method, as the needed accuracy can be a lot less to obtain similar results as with the traditional method.

$$\varepsilon_{max,calorimetric} = \varepsilon_{p_{out}} \cdot P_{out}/P_{loss}$$

Formula 6.9

Hence, in a typical 98% efficiency environment the measurement tolerance for the calorimetric losses can be 50 times higher than the measurement of the input and output power to obtain the same results.

7 Digital measurements

Power analyzers convert the input voltage and current signals from analog to digital signals and utilize digital signal processing techniques to calculate the quantities of interest, such as Average and RMS values, Power, Power Factor, Harmonics, and Efficiency. An example of a power analyzer measurement unit block diagram is shown in Fig 7.1

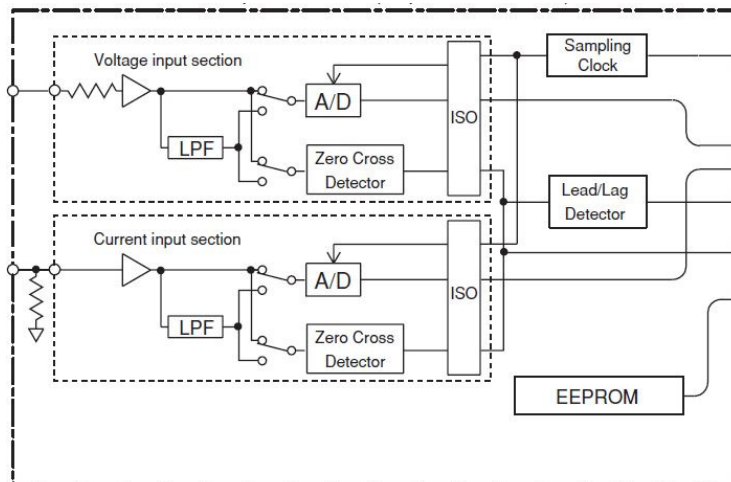


Fig 7.1 Power analyzer measurement unit block diagram example (Yokogawa WT230 manual)

Converting from analog to digital signals requires sampling the analog signal at a given discrete rate. The sampling frequency and the timing of the sampling in relation to the input signals are key factors in the measurement accuracy.

7.1 Sampling frequency

According to the Nyquist-Shannon sampling theorem, the sampling frequency f_s should be at least twice the highest frequency f_{in} contained in the signal so that the input signal can be accurately represented:

$$f_{in} \leq \frac{f_s}{2}$$

Formula 7.1

In practice, this means a high enough sampling frequency should be utilized, or the bandwidth of the input signal should be limited below the Nyquist frequency $f_s/2$. Aliasing effects occur if this condition is not met, i.e. when the power analyzer tries to measure frequencies above its specified range. Frequencies above the Nyquist frequency are “folded” down below the Nyquist frequency, resulting into an “alias” of the original signal and in measurement errors.

7.1.1 Oversampling

The input signal is oversampled when the sampling frequency is high enough to meet the condition indicated in Formula . Oversampling is illustrated in Fig 7.2

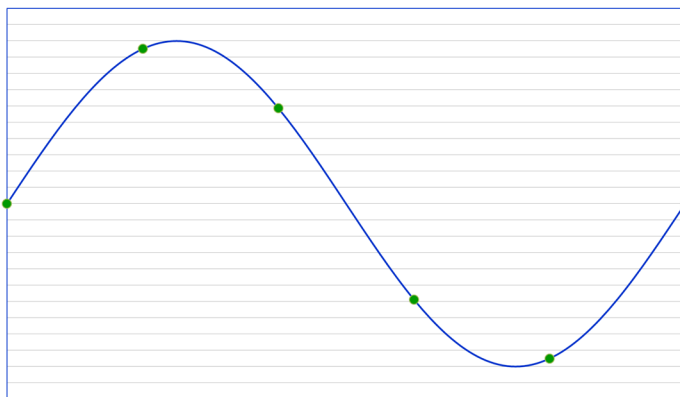


Fig 7.2 Oversampling $f_s = 5 \cdot f_{in}$

The higher the sampling frequency, the better the measurement accuracy is. Higher sampling frequency reduces the errors due to aliasing because the Nyquist frequency is pushed higher in the spectrum, where the frequency components of the measured signal tend to attenuate as the frequency increases. Oversampling gives an accurate representation of the input signal, which can be utilized to calculate Average, RMS, and Power values, as well as to reconstruct the signal in the time domain, filter it digitally, or make a FFT.

Switch-mode power supplies switch at frequencies typically above 50kHz, resulting in noise in the input and output lines, which could result in aliasing. Therefore, power analyzers give the possibility to introduce an anti-aliasing Low Pass Filter, as shown in Fig 7.1 The bandwidth of the filter should be low enough to attenuate the high-frequency noise, but high enough not to interfere with the typical mains distortions. For example, the bandwidth of the anti-aliasing filter is 500Hz in the Yokogawa WT230 power analyzer, or selectable 500Hz/5.5kHz/50kHz in Yokogawa's top of the line WT3000 power analyzer.

7.1.2 Undersampling

The input signal is undersampled when the sampling frequency is lower than twice the maximum frequency component of the signal, hence the condition indicated in 7.1 is not met.

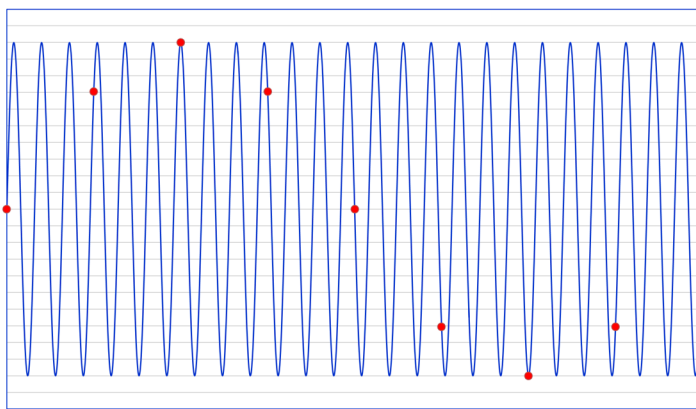


Fig 7.3 Undersampling $f_s = 0.32 \cdot f_{in}$

Undersampling results in aliasing and is illustrated in Fig 7.3. This means that sampled data cannot be utilized for analyzing the signal in the time domain, for digital filtering, or to make an FFT. However, undersampling can be utilized to calculate Average, RMS, and Power values, if the measurement is extended over several cycles of the input signal and assuming that the input signal is repetitive.

Let's consider that the sampling frequency f_s and the input signal frequency f_{in} satisfy the condition:

$$\frac{f_{in}}{f_s} = \frac{M}{N},$$

Formula 7.2

where M and N are integers representing the number of input signal cycles in the measurement period and the number of samples in the measurement period, respectively. The condition described by Formula 7.2 guarantees that an integer number of sample periods fit exactly within the measurement period. Additionally, if M and N are coprime, it is guaranteed that the same samples are not repeated over any two cycles of the input signal during the measurement period formula 7.2 can be written as:

$$\frac{f_{in}}{f_s} = K + \frac{P}{N},$$

Formula 7.3

where $K \geq 0$, $0 < P < N$, $N > 0$, and P and N are coprime. Usually N is an even number, $N = 2^n$ therefore P must be selected as an odd number. From Formula 7.3, we can see that one sample is obtained after every K periods of the input cycle and that after each sample period the sampling point advances with P/N over the input signal cycle. This means that we need to measure over $K \cdot N$ periods of the input signal to obtain N samples.

For example, if $f_{in} = 50\text{Hz}$, $K = 3$, $P = 1$, and $N = 8$, the sampling frequency should be $f_s = 16\text{Hz}$. This case is exemplified in Fig 7.3

7.2 Asynchronous and Synchronous sampling

Asynchronous sampling occurs when a non-integer number of input signal periods are sampled, as illustrated in 7.4. This may occur when the frequency of the input signal is not known or, for example, when the mains frequency deviates from the 50/60Hz nominal.

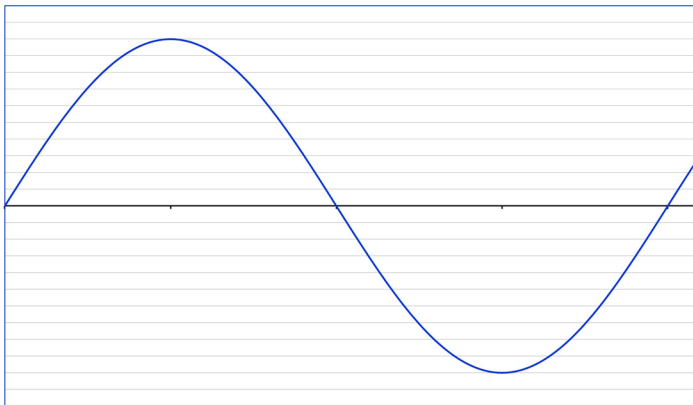


Fig 7.4 Asynchronous sampling

Asynchronous sampling introduces errors in Average, RMS, and Power calculations, as well as in harmonic amplitude and phase calculations.

The errors due to asynchronous sampling can be minimized with synchronous sampling, where the width of the sampling window is synchronized with an integer number of the input signal periods. This can be achieved by adjusting the sampling frequency to be an integer multiple of the fundamental frequency of the measured signal. For example, in the Yokogawa WT230 power analyzer, for input frequency $40\text{Hz} \leq f_{in} < 70\text{Hz}$, the sampling frequency is $f_s = 512 f_{in}$ and the sampling window width is 2 periods of the input signal.

7.2.1 Simultaneous and Non-simultaneous sampling

Simultaneous sampling refers to voltage and current signals being sampled at the same instant. This is essential for calculating the instantaneous power with high accuracy, as explained also in

section 4.3.3. Therefore, power analyzers have separate measurement channels and ADCs for simultaneous sampling of voltage and current signals, as illustrated in Fig 7.1.

Non-simultaneous sampling is utilized in systems with multiplexed input channels and one ADC that samples sequentially the signal from the input channels. Sequential sampling results in additional delay between the sampled voltage and current and is an additional source of errors.

Conclusions: Digital measurements are essential for high accuracy efficiency measurement.

Measurement errors can be minimized by sampling with high frequency, with proper filtering of the measured signals, synchronous sampling, and simultaneous sampling of voltage and current.

8 References

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