An universal high-voltage source based on a static frequency converter

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***Abstract*: For testing electrical equipment the test voltage has to achieve special requirements, which are noted in the international standards. Tests showed that it's not possible to fulfill those requirements with a passive pulse pattern. This paper deals with the development of an electronic control unit for a static frequency converter. The focus is on the synthesis of the control and the obtained results after realizing in a static frequency converter with semiconductor valves.**

I. Introduction

IEC 60060-1 defines HV test voltages for common use, but in addition there is the demand of HV tests with high sinusoidal AC voltages of higher or lower frequency than power frequency or even with non-sinusoidal shape. This is important for the insulation research for future power systems, but even today some tests have to be performed at higher frequency, e. g. at test objects with magnetic saturation effects. Alternating voltages of a wide variety of voltage shapes can be generated by static frequency converters if the converter is controlled in an appropriate way. In addition to the voltage shape, there are many other requirements, e.g. a low PD level caused by the converter.

Let us assume an AC voltage of higher frequency shall be generated by a frequency converter. To achieve a high voltage a step-up transformer is required. At the output of this transformer the sinusoidal voltage is distorted with a multitude of harmonics. Responsible for this is the pulse pattern which can't create a continuous voltage. Due to this, the voltage has to be smoothed with a reactance or a multistage sinus filter. However, this would result in a huge power loss for higher power ratings of the static frequency converter. On the other hand the static pulse pattern does not adapt to different loads and thus the total harmonic distortion for a sinusoidal voltage is higher than 5% (IEC 60060-1).

To prevent these issues a novel control strategy for static frequency converters was developed to meet the requirements. This paper deals with the automatic control system which provides a total harmonic distortion factor smaller than 5% during testing with up to 30kVA output power. The control system is based on a comparison between actual and nominal voltage. The difference of these voltages is processed in a regulation with a proportional and integral element. Beyond this the actual current is measured and processed for an inner servo loop. Taking into account that the operation of power semiconductors is constrained by different time limits and that forbidden inverter-states need to be interlocked, Programmable Logic Devices (PLDs) are inserted between the regulation output and the Insulated Gate Bipolar Transistors (IGBTs). Thus the pulse pattern does not remain static but is dynamically adapted to the ideal voltage form in real time.

The principle arrangement of the considered converter with passive LC-filter is shown in Fig. 1.

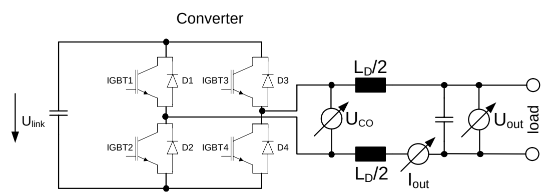

Fig. 1: Frequency converter

Beside the mentioned sinusoidal frequency variable AC voltage it's also possible to generate such of triangular or nearly rectangular shapes with the assistance of the new control.

II. Control Synthesis

There are many possibilities to regulate the output voltage. It's possible to control the well known pulse-width modulation by variation of the control voltage. However simulations show that this control is too slow and due to this there is no significant improvement of the output voltage. Another attempt is the direct control, which triggers the valves of the converter directly. This principle offers a multitude of possible realizations, which have to be examined carefully. The intention of the synthesis is to develop an appropriate closed-loop control structure and a control element, which both are stable and stationary correct, sufficient damped, but also quick enough.

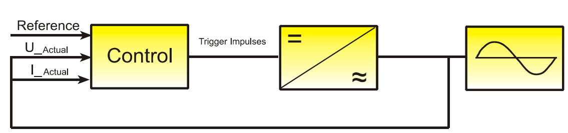

Fig. 2: Basic concept for the control

The basic idea for the control is based on a simple setpoint / actual value comparison applied to the output voltage and the output current. However, if one realizes this option the converter would permanently seesaw at the output between $+U_{link}$ for $U_{ref} < U_{actual}$ and $-U_{link}$ for $U_{ref} > U_{actual}$. This easy control would not be stationary correct as well as tend to oscillations and instabilities. Due to this a third possible state has to be implemented: the free-wheel-state, at which the output of the converter is 0 V. Therefore a tolerance zone around the reference voltage has to be defined. For the enhancement of the attributes for the closed-loop control, different control structures were examined.

During the simulation the cascade control, shown in Fig. 3, was proved as the best control conception. Here the control of the output current is superimposed by the control of the output voltage.

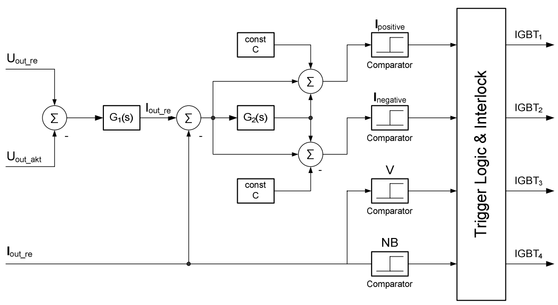

Fig. 3: Cascade control for one phase

In Fig. 4 the cascade control, the converter, LC-filter and the load are displayed in a control engineering block diagram.

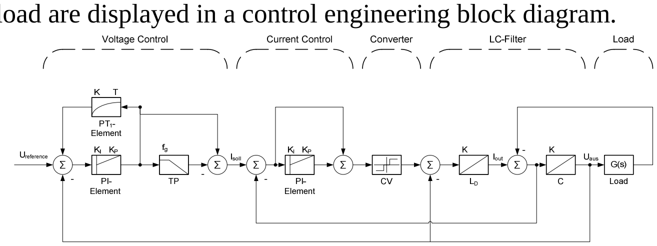

Fig. 4: Control engineering block diagram of the cascade control

A. Current Control

Designing the current control some facts have to be considered:

- Inductance of the reactor L_D
- Voltage of the link U_{link}
- Output voltage (amplitude) U_{out}
- Size of the free-wheel
- PI-control parameters
- Converter switching rate

The following paragraphs explain these facts in detail.

The ripple of the output current is very dependent on the inductance of the reactor L_D. Due to the used power semiconductors the minimal period T_p is less than 100µs. The minimum condition time must be less than 5µs. Regarding only (1) with I_{out} the output current and U_{link} the voltage of the link it's obvious that the best choice to minimize the current peaks would be a very big inductance.

$$I_{out} = \frac{1}{L_D} \int U_{link}\, dt \tag{1}$$

However large voltage drops along the inductances would occur. As a result of these the usable voltage at the output U_{Out} for the complete converter is reduced. Even if one reduces the setpoint value of the output voltage the current peaks remain the same order and the ripple rises heavily. Beside the bigger voltage drop an exalted inductance has a negative impact on the dynamic behavior of the whole system. With (1) the rise time t_{rise} is

$$t_{rise} = \frac{I_{out_n} \cdot L_D}{U_{link}}\,. \tag{2}$$

With formula (2) the rise time t_{rise} is calculated to 100µs up to 350µs for the different values of possible inductance. Little inductance on the other hand causes high overshoots and a long settling time, which does not occur for greater inductances. At this point the well known problem of the automatic control engineering appears. A compromise between enough speed and an adequate damping has to be found. The dynamic performance of the system has to be good enough to obtain the desired steepness of the output current.

Another important property of the current control is the width of the free-wheel band. A narrow free-wheel band stands for little tolerance between the setpoint reference current $I_{out,ref}$ and its actual value. In this case the power flow toggles between feeding the load and energy recovery. Free-wheel, as a possible third state, is not used. Output voltage either is $+U_{link}$ or $-U_{link}$. The pulse pattern of the converter the reference and actual value of the current is shown in Fig. 5. A variation of the constant K, representing the width of the free-wheel band, discovers the optimum value. Comparing Fig. 5 and Fig. 6 the change in pulse pattern is more than obvious: For the optimum the three possible inverter states feed, energy recover and free-wheel are used.

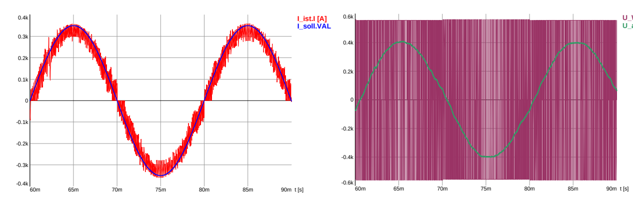

Fig. 5: Current control with a small free-wheel

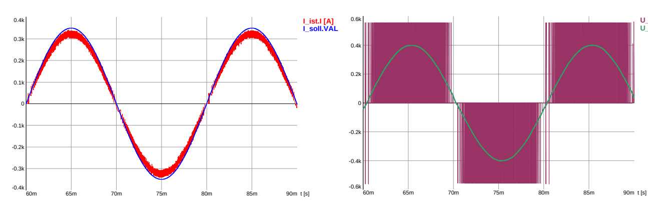

Fig. 6: Current control with an optimum free-wheel band

The inverter's pulse pattern "settles" and the appearing pulses at the output voltage get smaller, as long as the difference between actual and nominal value remains sufficiently small. For a bigger value the actual current follows its nominal value insufficiently, especially in the crest area.

This simulation was done only for a purely non-inductive load. For an inductive load the previously discussed behavior of the inverter states is still the same. Merely the state of energy recovery from the load is much more important and more frequently in use. The simulation results also show, that if the free wheel band is properly attuned a toggle between $+U_{ZK}$ to $-U_{ZK}$

can be avoided and the free-wheel state is used. Due to this the pattern recovery – free-wheel – recovery or feed – free-wheel – feed and so on is generated.

A bigger free-wheel band leads to a breakdown of the output current near the zero point of the desired value curve. To prevent this problem an additional PI-element was integrated in the current control system (see Fig. **4**). This ensures that the average value of the current follows its specified value. Because the dynamic performance and therefore the stability of the system is not allowed to be endangered, the time constant of the I-element has to be longer than the switching period of the converter. To balance possible direct components in the output current the time constant of the integrator is set to the period time of the output current: T_I=20ms or K_I=20. The direct component can occur due to little deviation of the absolute values of the free-wheel constants +K and –K or asymmetries in the comparator thresholds. Therewith the actual current only follows the period average of the desired value. This means, that the output current does not have any DC components. Because of the reduction of the factor K_I the current control does not exactly follow its desired value. A system deviation remains that means the current control does not have a steady-state behavior. However, this is not necessary because the current control works subordinated. The superposed voltage regulation is responsible for the steady-state accuracy of the complete system.

B. Voltage Control

Apart from the few discussed exceptions the considerations which are made so far lead to the assumption that the actual current value follows its desired value. In the following section they are treated as one identical value $I_{out,act} = I_{out,ref}$. Furthermore the current control is considered as a steady state system. With the aid of these assumptions the combination of the voltage control simplifies to a complete linear time-invariant system (Fig. 7).

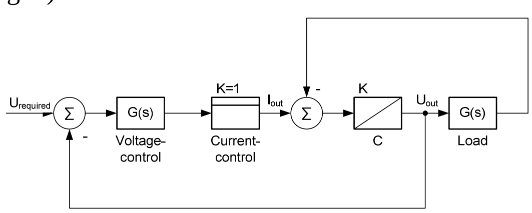

Fig. 7: Block diagram for the voltage control

In the block diagram the assumed optimum current control is reproduced as a proportional element with the amplification K = 1. This substitutes the PI-element, the three-step element (converter) and the I-element (reactance induction L_D) from Fig. 4.

For the design of the voltage control the system has to be simplified in a second step. Only the transfer function of the controlled system and the control itself remain as components of the presented system. The result is shown in Fig. 8 with

$$G_{St}(s) = \frac{1}{C \cdot s} \Big/ 1 + \frac{1}{C \cdot s} \cdot G_{load}(s)\,. \tag{3}$$

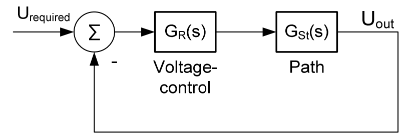

Fig. 8: Simplified block-diagram of the complete system

For all load cases the main focus for the control is to compensate the habits of the path $G_{St}(s)$. Thus the control is depending on the load and the design has to be made with respect to the different loads. The first step is to develop a basic control assuming as the easiest case, an ohmic load. With

$$G_{Load}(s) = \frac{1}{R_L} \tag{4}$$

is

$$G_{St}(s) = \frac{R_L}{1 + C \cdot R_L \cdot s}\,. \tag{5}$$

To fulfill the steady-state accuracy an I-element has to be implemented in the regulation. With a proportional element the dynamic should be hold up whereby you get the following sum

$$G_R(s) = \frac{K_I}{s} + K_P \tag{6}$$

$$\Rightarrow G_R(s) = K_{PI}\,\frac{1 + T_{PI} \cdot s}{s},\; K_{PI} = K_I,\; T_{PI} = \frac{K_P}{K_I}\,. \tag{7}$$

With the aid of the Nyquist criteria the system stability was proved.

The ohmic-capacitive load showed an equivalent behaviour and further investigations are not necessary.

For an ohmic-inductive load the situation is different. If only a PI-control is in use, the system does not react on a load with a negligible ohmic part. Direct current component in the output current entirely flow through the inductances and do not contribute to the voltage drop over the output capacitance with which the actual value for the voltage control is measured. However, real measurement systems always have a small offset and thus a direct current component. This DC component in the measurement signal is leading to a steady rise of the output signal. To prevent this problem an additional PT_1-element negative reverse onto the input has to be implemented. So the ramp up can be prevented, but the output current still has a constant direct current component due to the amplification of the PT_1 element. With the aid of a low-pass filter it's possible to determine the low frequency components and subtract them from the original signal. Fig. 9 shows the voltage control and the second order filter which is made of two PT_1 elements.

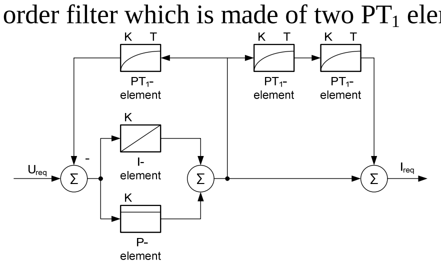

Fig. 9: Block diagram of the voltage control with PT1- feedback and filter

III. RESULTS

After developing this control the next step was to simulate the whole system consisting of a DC source for the converter link and the converter itself controlled by the new closed loop regulation. The simulation results with different loads were very good and thus an already existing single phase pulse-width modulated converter was modified and fit out with the new regulations. Tests were carried out with all possible combinations out of ohmic, inductive and capacitive loads. The maximum tested output power was 30kVA. Table 1 shows the summery of the results.

TABLE I
RESULTS FOR THE MEASURED OUTPUT VOLTAGE

Load	2,3 Ω	2 mH	2,3 Ω + 10 µH
Voltage	200 V	134 V	340 V
Current	83 A	215 A	69 A
THD	0,5 %	3,0 %	1,2 %
Fig. No.	10	11	12

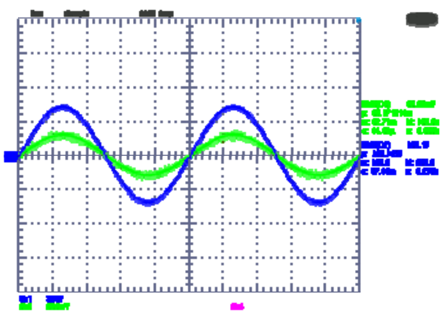

Fig. 10: Oscilloscope screenshots for ohmic loads (voltage = blue, current = green)

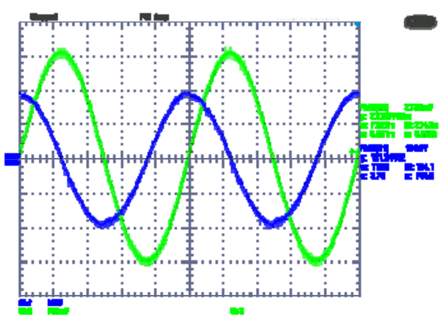

Fig. 11: Oscilloscope screenshots for inductive loads (voltage = blue, current = green)

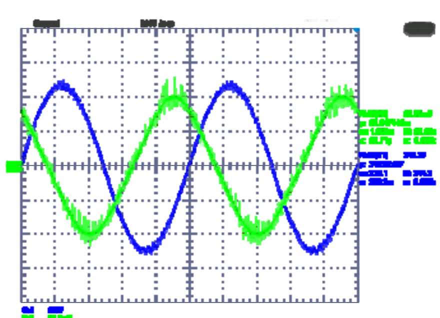

Fig. 12: Oscilloscope screenshots for ohmic-inductive (voltage = blue, current = green)

Fig. 13 shows the variety of the closed-loop control. It's possible for the output voltage to follow any reference value. Even for a triangle like Fig. 13 or a rectangle which is not displayed.

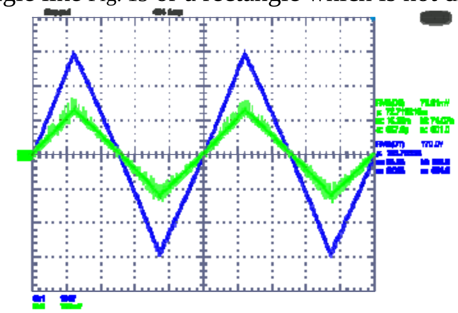

Fig. 13: Oscilloscope screenshots for different loads (voltage = blue, current = green)

IV. CONCLUSION

After the design of the closed-loop control and a successful simulation of the complete system, consisting of the power line, rectifier, link, alternating current converter and different loads the hardware was developed to implement the control in a real system. With the aid of this development an existing pulse-width modulated, single phase rectifier was modified to a new regulated one.

Measurements during operation of the modified converter shows that for the sinusoidal AC voltage a total harmonic distortion factor (THD) less than 5% can be obtained. Thus the standard requirements can be fulfilled with the real setup.

The control is even so variable in reality that apart from the sinusoidal it's possible to generate any required pulse form for an ohmic load.

The direct control of the IGBTs, which was developed for a frequency variable generation of test voltage, guarantees testing with output voltages conform to international standards for all the loads that were examined. An adaptation of the control parameters to different loads is hardly necessary. By using the better aligned and finer pulse pattern it is possible to smooth the voltage output with a much smaller inductance inline and a capacitor parallel to the phases. This reduces the losses and increases the output power rating for the whole system.

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