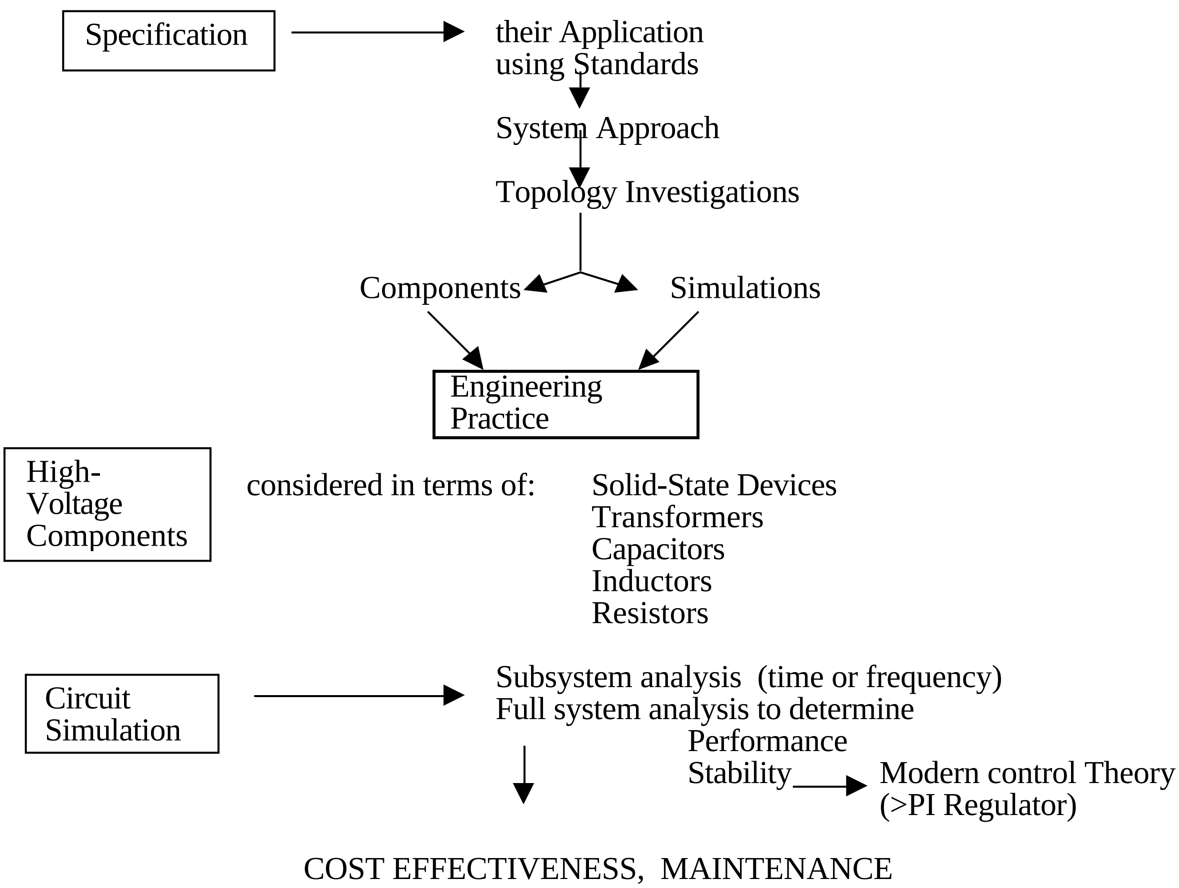
Session 1-4
Chairman : H.U. Boksberger, PSI

High Voltage and Energy Storage

REVIEW OF SESSION 1.4 - HIGH VOLTAGE AND ENERGY STORAGE

Hans U. Boks berger (Chairman)
PSI

This session looked high voltage power supply design and digital regulation systems for precise control. There was also an interesting paper that led to reflections on storage capacitor design for high-power, high-voltage networks, such as PFNs in line-type modulators. Some first results of tests of a polyphase boost-converter-modulator were also presented. All of these papers had a common theme of requiring a dedicated and structured approach to the system design using these high-power elements.



POWER SUPPLIES FOR TESLA MODULATORS

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Abstract
Modulators are used to generate the pulsed power for the klystrons of the superconducting linear accelerator TESLA. They produce rectangular high voltage pulses of up to 120 kV. The electrical power during the pulse is typically 15 MW and can maximally be 16.8 MW. The pulse length is 1.6 ms with a repetition rate of 5 Hz, for app. 10 % of the modulators it is 10 Hz. This leads to a needed pulsed power of 8.9 GW. It is obvious that this energy can not be taken from the mains directly. Therefore it is stored in capacitor banks to be released during the pulse. Power supplies are needed to recharge the capacitor banks and to decouple the low repetition rate from the mains. The electrical supply companies have very strict rules of the amount of distortions that are allowed to be produced by a customer especially in the frequency range below 25 Hz (flicker frequencies). To meet these rules the power supplies have to operate in constant power mode. Different types of power supplies have been investigated to check a possible use for TESLA

1. INTRODUCTION

TESLA klystrons require high voltage pulses of up to 120 kV with a pulse power of up to 16.8 MW. The pulse length is 1.6 ms with a repetition rate of 5 Hz, 10 % of the klystrons are working a 10 Hz repetition rate. The pulses are generated in modulators. In order not to take the pulsed energy from the mains these modulators store energy which is released during the pulse. The energy storage is then constantly loaded from the mains. To do so different types of power supplies have been investigated for the use in the modulators.

2. BOUNCER MODULATOR

Beside other designs the bouncer modulator is the most promising solution for the modulators in respect to cost and ease of design. Fig. 1 shows the principle schematic with the main capacitor bank, a semiconductor switch, the pulse transformer, the bouncer circuit and the HV power supply. For safety reasons ignitrons are installed to quick discharge the main capacitor bank in case of failure in the klystron.

To generate the HV pulses the main capacitor bank is charged to a voltage at the 10 kV level. Via the semiconductor switch the pulse transformer is connected to the capacitor bank. With the step up ratio of 1:12 the voltage is transformed to the 120 kV level.

During the pulse the voltage of the main capacitor droops for about 19 %. The principle can be seen in Fig. 2. To correct the voltage droop during the pulse to +/- 0.5% a bouncer circuit is used. This is a resonant LC circuit which creates a low frequency sine wave which is triggered slightly before the main pulse. The main pulse is positioned in the linear part of this sine wave. The sum of both voltages is a rectangular voltage of the desired parameters. With the use of the bouncer the stored energy inside the modulator is decreased.

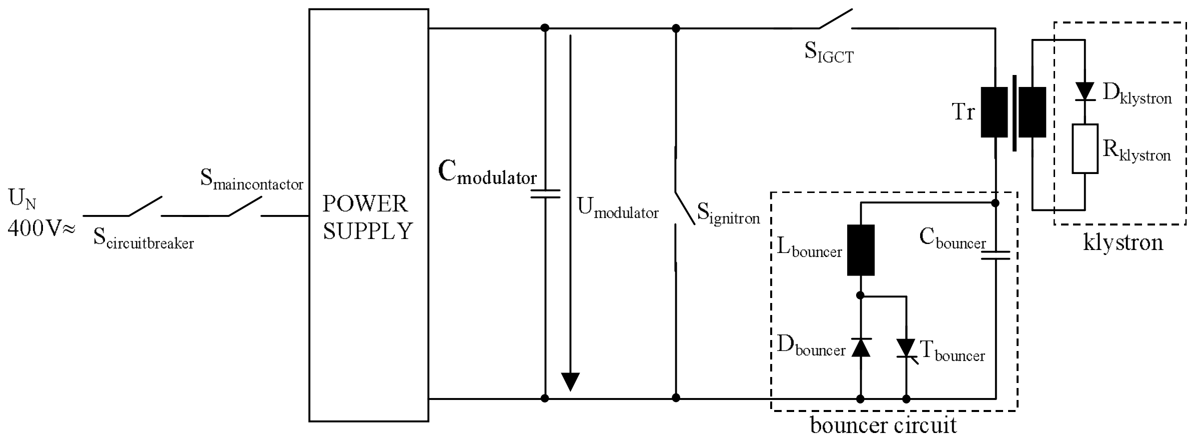


Fig.1: Principle schematic of the bouncer modulator

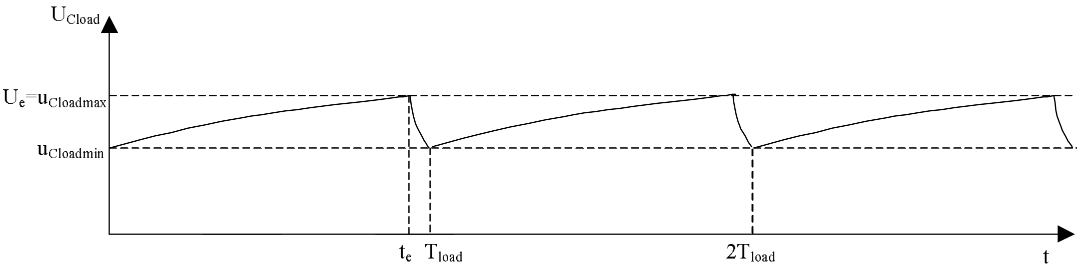


Fig. 2: Voltage curve form of the main capacitor bank in the modulator

3. DISTURBANCES TO THE MAINS

3.1 Pulsed power of the modulators

In TESLA 584 modulators will be in operation. Another 12 modulators are in stand by mode ready to start operation in case of a failure of the working units. 519 modulators operate with the 5 Hz repetition rate. The other 65 modulators will have a repetition rate of 10 Hz. This leads to a typical total peak pulse power of 8.9 GW at 5 Hz repetition rate and 1 GW for the intermediate pulses at 10 Hz repetition rate. This power is taken during 1.6 ms. During 200 ms respecting 100 ms the energy taken from the capacitors has to be recharged without disturbing the mains.

3.2 Allowed distortions to the mains

The German standard VDE 0838 or the equivalent European standard EN 61000 defines the amount of distortions that are allowed to be produced by a consumer of electrical energy. These distortions are defined as relative voltage changes d with

$$d = \frac{\Delta U}{U} \approx \frac{\Delta S}{S_{sc}}$$

d = allowed relative voltage changes

U= mains voltage

ΔU= variation of mains voltage

ΔS = variation of power due to the modulators

S_{sc} = short circuit power of the mains

In general no consumer of electrical energy is allowed to produce more than 3 % of voltage variation to the public mains. For low repetitive changes below 25 Hz this value is even more decreased. Voltage changes below 25 Hz are seen as changes in the luminance of the electric light. Since the human eye is very sensitive to these changes this appears as flickering light. These frequencies are called flicker frequencies. The value d can be achieved from diagrams in the standards. For the 5 Hz repetition rate the value d is < 0.5 %.[3,8]

TESLA will have a distributed electrical power supply system with a voltage of 20 kV. With the short circuit power S_{sc} of app. 200MVA per service hall the allowed power variation can be calculated to:

$$\Delta S < 200MVA * 0,005 = 1MVA$$

In each service hall up to 100 modulators are to be installed. The typical real power consumption is assumed to 15 MW per hall. Therefore each modulator shall not produce more the 10 kVA variation over the entire time between the pulses. With a nominal input power of 150 kW this leads to an allowed variation of 6.5 % of this value including the reactive power changes.

For the 10 Hz operation the curve of relative voltage changes has a minimum. The allowed voltage variation is decreased to d = 0.25 %. The number of modulators working at the 10 Hz level is low. There are 55 modulators installed in the service hall on the DESY site. The power supplies have an allowed variation of 3 % of their nominal power. The remaining 10 power supplies are installed in another service hall having a separate mains connection point.

4. TYPES OF POWER SUPPLIES

4.1 General assumption

There will be one power supply for each modulator. It will have a standard 400 V three phase input. The output voltage will be 12 kV_{DC}. The nominal power will be 150 kW for the 5 Hz and 300 kW for the 10 Hz repetition rate. The typical power needed for the 5 Hz operation is 120 kW. The power supplies will be built in modules. By this a high reliability and a good maintainability is given.

The advantages of having one power supply per modulator are:

- very high redundancy in the rf system. A failure of a modulator or a power supply does not affect any other modulator
- a failure in a single power supply module will not turn down the modulator
- each power supply can be regulated independently with a high regulation dynamic
- at the low voltage level switch gear is available as low price commercial of the shelf component.
- in case of replacing or working at the power supplies no further high voltage safety requirements are given

During operation the power supply has to meet two requirements:

- The main capacitor bank has to be recharged to an accurate value of voltage in order to obtain the same voltage at the klystron from pulse to pulse. The accuracy has to be +/- 0.5 %.
- The low repetition rates of 5 Hz and 10 Hz have to be suppressed in order not to produce disturbances to the mains.

4.2 Topologies of the power supplies

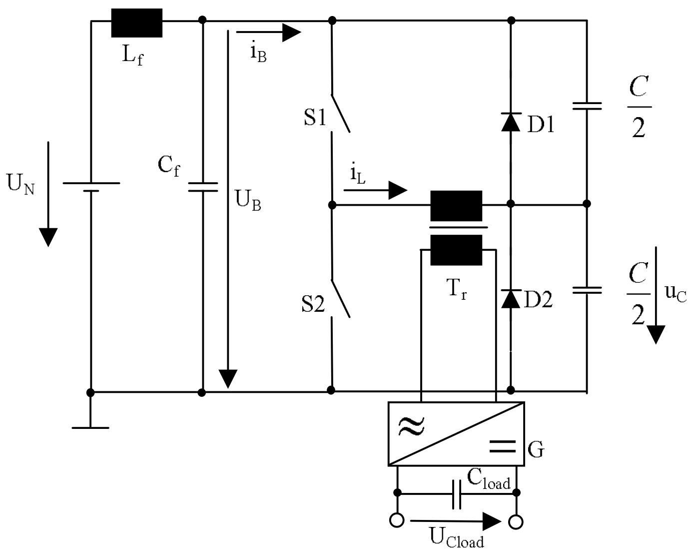
The given definition of the allowed distortions to the mains restricts the topologies of power supplies. The power supplies used in the first prototypes of the modulators at TESLA Test Facility are primary regulated SCR bridges with transformer and secondary diode bridge. This can not be used for TESLA due to the large variation in reactive power. Three phase rectifiers using SCRs like B6 bridges cannot be used for the same reason. Switched mode power supplies in the needed power and voltage range have recently come to the market and are available from industry. Additionally a digital regulation has to be introduced to fulfill the requirement for the constant power consumption.

Among other solutions described in [1,8] the following topologies have been investigated at DESY.

- Series resonance converter
- Series connection of buck converters
- Hybrid power supply
- SCR bridges in series with a diode rectifier

4.3 Series resonance converter

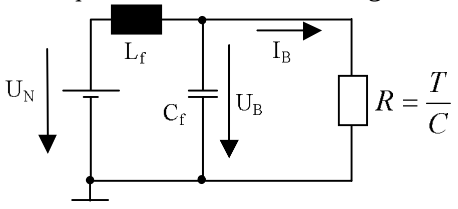
The power supply shown in Fig 3 was developed at DESY. This topology is known for small power supplies for auxiliary voltages. So far it has not been used as power supply to charge large capacitive loads in a constant power mode.



G	Rectifier
i_B	supply current
i_L	primary current of the transformer
u_C	voltage of the resonance capacitor
U_{Cload}	output voltage to the switch of the klystron
i_{Bt1}	current i_B at the time t1
L	primary stray inductivity of the transformer
U_B	supply voltage
f	resonance frequency of the resonant circuit of L and C
ü	gear ratio of the transformer and rectifier
T	period time of the switching frequency of S1 and S2
C	resonance capacitor
U_B	supply voltage
U_N	line voltage

Fig. 3: Series resonance sine converter

The equivalent circuit according to the arithmetic average of the supply current I_B is shown in Fig.4:



T	period time of the switching frequency of S1 and S2
C	capacity of the resonance capacitor
R	equivalent resistance of the switch mode power supply
U_B	supply voltage
U_N	line voltage
I_B	average supply current i_B

Fig 4: Equivalent circuit of the power supply

The equivalent circuit of the switch mode power supply is a resistor R which is constant when the period time T of the switching frequency f (10 kHz– 20 kHz) is constant. This resistance is independent of the capacitor voltage U_{Cload} and the pulse repetition rate of the modulator. By this the input power is (in a wide range) independent from the voltage of the main capacitor bank of the modulator.

4.4 Buck converter

To achieve the high voltage level of the main capacitor bank it is possible to stack a group of low voltage power supplies. This topology is used for power supplies that are manufactured in industry. These units are now under construction and will be used for some of the TTF modulators. The power supplies are buck converters having a nominal voltage of app. 750V each. 16 modules are stacked to deliver the required voltage of 12 kV. The transformer has a 400 V input and 16 outputs of which 8 outputs are in delta, the other 8 outputs are in star to get less mains harmonics. Fig. 5 shows the principle diagram of such a power supply.

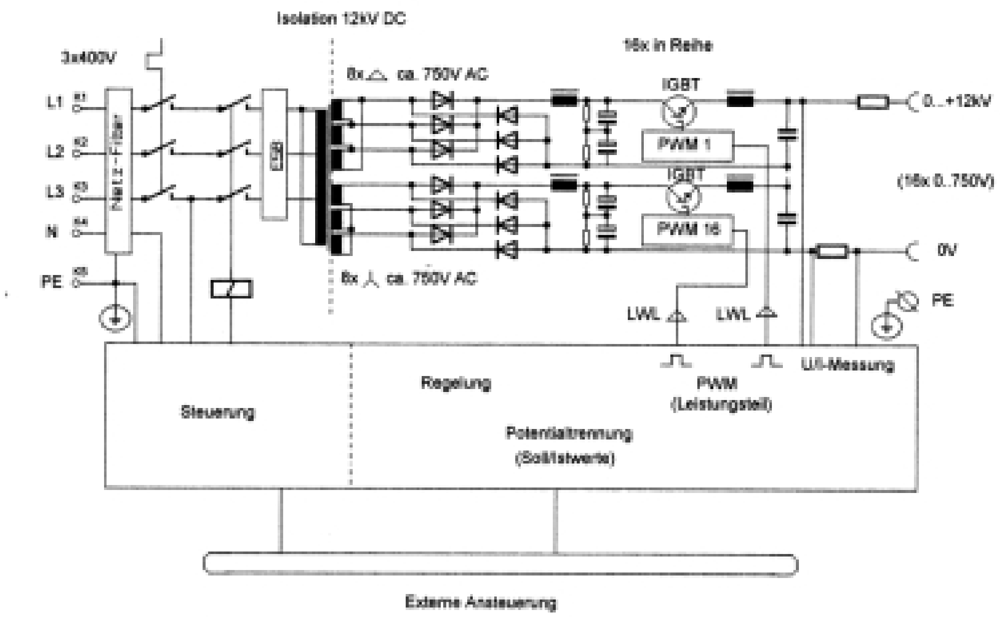


Fig. 5: Power supplies with stacked buck converters

The buck converters are working with a Pulse Width Modulation (PWM). The overlaid regulation for the voltage and the constant input power is accomplished on an external board that is developed and provided by DESY. The internal regulation of the power supply is a power regulation. The units will receive the power reference signal from the DESY regulation and transform this into voltage and current values.

4.5 Hybrid power supply

When looking at the development of the cost of switched mode power supplies a permanent decrease can be seen. In the last 5 to 10 years the prices have been reduced by a factor of two. Nevertheless they are still high in comparison with SCR or diode technology. Therefore another solution is considered. This is the hybrid power supply.(Fig 6)

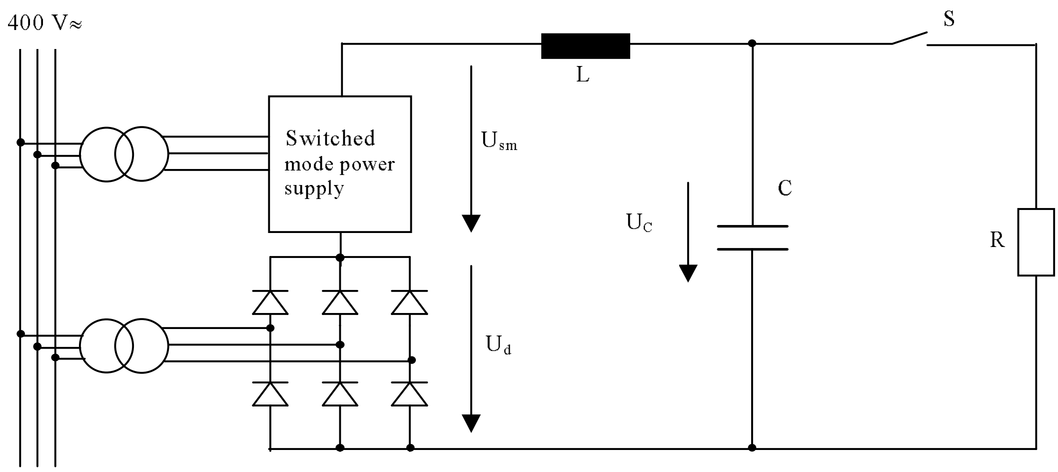


Fig. 6: Hybrid power supply

Basic idea

When looking at the waveform of the capacitor bank voltage of the modulator (Fig. 2) this can be splitted into two parts. There is a constant DC part and a part of varying voltage. The basic idea is to produce the constant DC part with an unregulated diode rectifier. The changing part will be produced by a switched mode power supply e.g. a buck converter. By this combination the full regulation dynamic of the switched mode supply can be used. Since the price of the diode rectifier is app. 40 % to 50 % of the price of a switched mode supply, an overall price reduction of 20 % to 30 % seems possible depending on the chosen value of the DC voltage. The principle is proven in simulations.

4.6 SCR supply with diode rectifier

The same basic idea as for the hybrid supply is assumed. Here the switched mode supply shall be replaced by SCR technology to further reduce cost. The voltage of the diode rectifier is at 10 kV. The SCR has to be in inverter mode at the beginning of the loading period to decrease the voltage. At the end of the loading period the power supply has to add voltage. A principle schematic is shown in Fig. 7. The voltage curves are shown in Fig. 8.

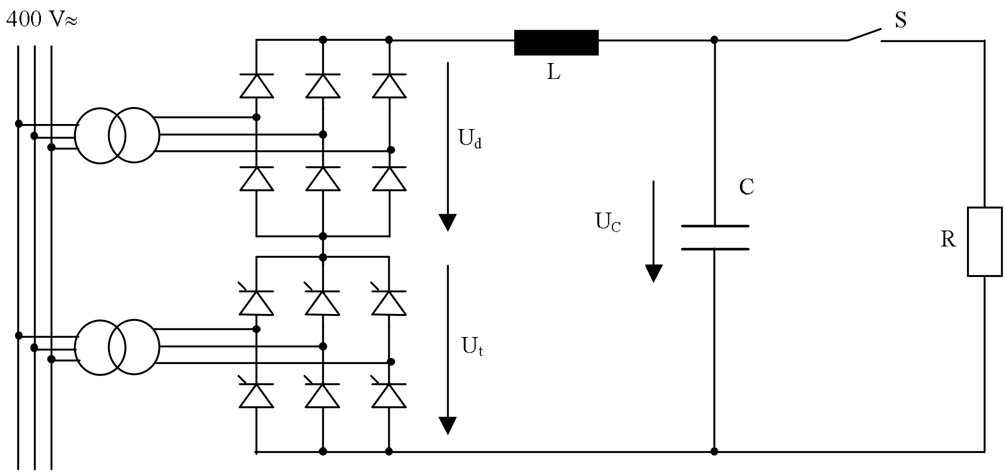


Fig.7: Series connection of a diode and a SCR bridge

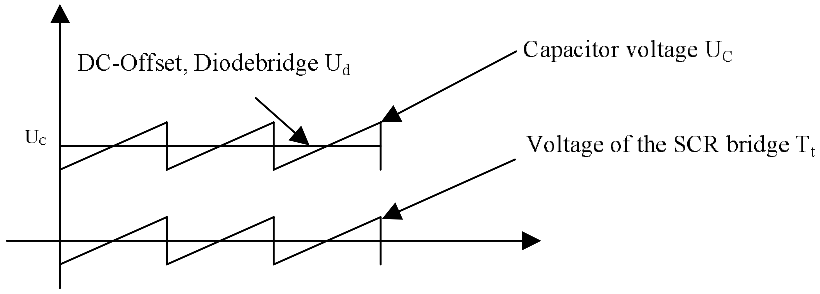


Fig. 8: Voltage curve of series connection of diode and SCR bridge

Here two versions are possible:

- The SCR power supply has a single phase angle control
- The SCR power supply is driven in sequential phase control

Both solutions have to take care for the reactive power of the SCR supply in order to keep the distortions of the mains low.

5. CONCLUSION

The low repetition rate of TESLA with the high pulsed power lead to hard specifications for the power supplies. Different types of power supplies have been investigated in terms of functionality and price. The result is that there are good solutions. The most convenient type with respect to regulation dynamics is the switched mode power supply but the price is still high. A good compromise is the hybrid power supply, a series connection of a diode bridge and a switched mode supply, since low price and good regulation abilities are combined. The third topology is the series connection of a diode bridge and a SCR rectifier. The drawback here is the large filter choke that has to be introduced to smooth the DC current. The regulation for this power supply is difficult.

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DIGITAL REGULATION FOR TESLA MODULATOR POWER SUPPLIES

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Abstract

Modulators are used to supply pulsed power to the TESLA klystrons. Inside the modulator a capacitor bank stores energy to be released during the pulse. During the pulse the voltage droops for about 19 %. The capacitor is charged by a special power supply which suppresses the changes of power consumption from the mains. For these power supplies a regulation was developed. This regulation keeps the input power of the power supply constant. The accuracy of the initial klystron voltage from pulse to pulse is better than 0.5 %. A digital regulator based on a programmable ALTERA device including a RAM, a pre- filter, a linearisation and a self- learning algorithm was built. Via VME interface a communication with a control computer is possible. With an adequate pre- filter it is possible to regulate different kinds of power supplies which are used in modulators.

1. INTRODUCTION

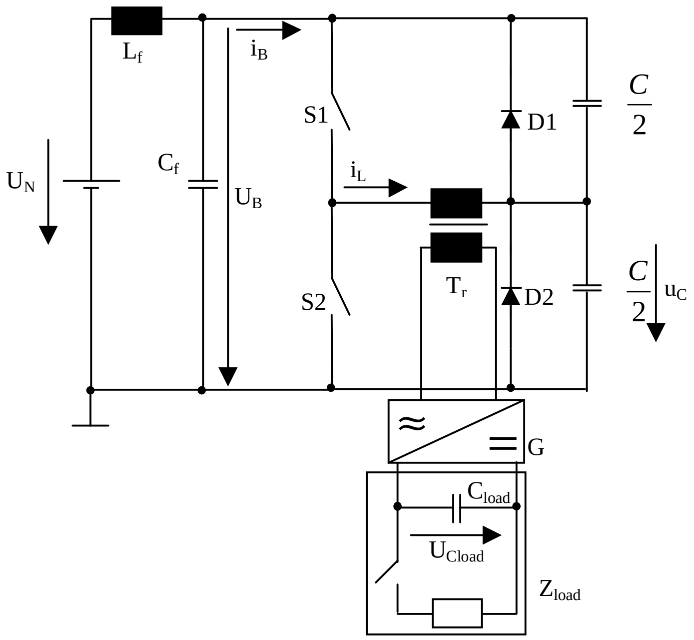
To suppress the pulsed power consumption of klystron modulators from the mains two solutions have been developed at DESY.

1. A power supply with system specific constant power consumption. It is possible to use the power supply with a simple regulator including a pre- filter.
2. A digitally self learning regulator. With different pre- filters this regulator is able to work with many types of power supplies.

To take advantage of both solutions the special power supply and the digital regulator are combined in one modulator at DESY.

2. SERIES RESONANCE CONVERTER

The power supply shown in Figure 1 was developed at DESY.

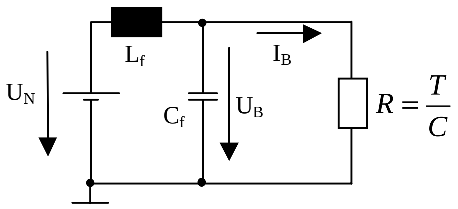


G	Rectifier
i_B	supply current
i_L	primary current of the transformer
u_C	voltage of the resonance capacitor
U_{Cload}	output voltage to the switch of the klystron
i_{Bt1}	current i_B at the time t1
L	primary stray inductivity of the transformer
U_B	supply voltage
f	resonance frequency of the resonant circuit of L and C
ü	gear ratio of the transformer and rectifier
T	period time of the switching frequency of S1 and S2
C	resonance capacitor
U_B	supply voltage
U_N	line voltage
Z_{load}	the load at the output may be any electrical circuit, eg. a capacitor, a switch and a klystron

Figure 1: Series resonance converter

The capacitor C_{load} stores the energy for the klystron pulses. The zero current switching power supply supplies constant power to the load Z_{load} when the switching frequency f of S_1 , S_2 is constant.

The equivalent circuit according to the arithmetic average of the supply current I_B:



T	period time of the switching frequency of S1 and S2
C	capacity of the resonance capacitor
R	equivalent resistance of the switch mode power supply
U_B	supply voltage
U_N	line voltage
I_B	average supply current i_B

Figure 2: Equivalent circuit of the power supply

The equivalent circuit of the switch mode power supply is a resistor R which is constant when the period time T of the switching frequency f (10 kHz – 20 kHz) is constant. This resistance R is independent of the capacitor voltage U_{Cload} and the pulse repetition rate of the modulator. Therefore the input power is (in a wide range) independent from the voltage of the main capacitor bank of the modulator.

To describe the function of the power supply Figure 3 is used.

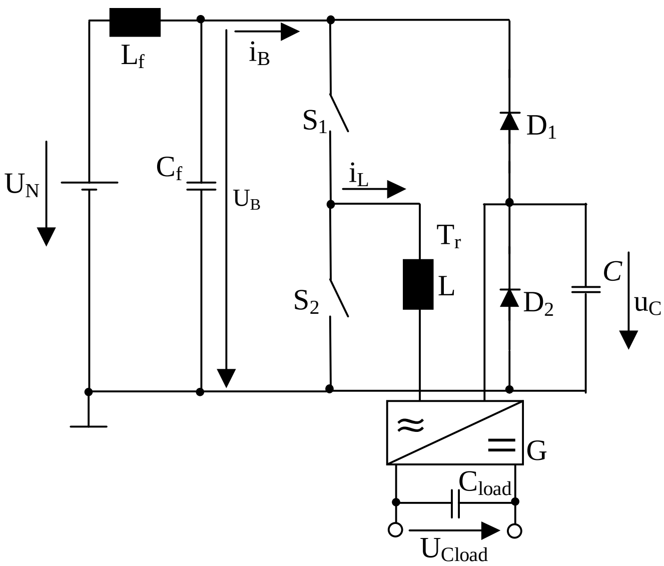


Figure 3: Series resonance converter with only one capacitor

The corresponding voltage and current waveforms of the following explanation may be seen in Figure 4.

Assumption: u_c=0V, i_L=0A and S_1 turning on

Because of the stray inductance L of the transformer the current i_L keeps zero while turning S_1 on (zero current switching power supply). Then the current i_L raises and the resonance capacitor C is charged. While charging the resonance capacitor C the supply current i_B is equivalent to the resonance capacitor current i_L.

When the resonance capacitor voltage u_C reaches the supply voltage U_B diode D_1 will start to conduct. The current i_L continues flowing forced by the energy stored in the stray inductance L of the transformer. The energy stored in the stray inductance L of the transformer will slowly be passed to the load C_{load} and the current i_L decreases linearly to zero.

After that a new cycle can start by turning S_1 off and S_2 on. With each loading and unloading period the same amount of energy is transmitted from the primary side of the transformer to the main capacitor bank. This amount is the energy of the resonance capacitor loaded to the voltage U_B or unloaded to 0.

In Figure 1 the resonance capacitor C is divided into two capacitors $\frac{C}{2}$ having half the capacitance of C. The same principle as for the half bridge with one resonance capacitor C is valid. The difference is that the frequency of the current i_B is doubled and the amplitude of i_B is divided by two . When one of the capacitors is loaded, the second capacitor is unloading and vice versa.

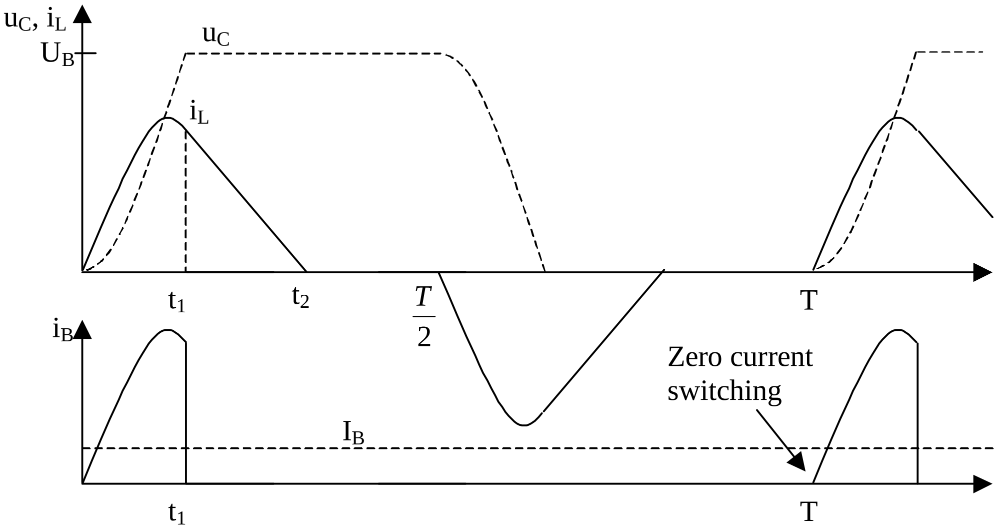


Figure 4: Voltage and current functions of the half-bridge

Please note that the current pulses are in the range of 20 kHz. They are filtered by the input filter and do not pass to the mains.

Derivation of the arithmetic average supply current I_B:

While charging the resonance capacitor C the supply current i_B is equivalent to the resonance capacitor current i_L. Therefore the charge Q taken from the power supply to the resonance capacitor C equals $Q = CU_B$.

The charge Q can be expressed as follows:

$$Q = \int_0^T i_B \, \mathrm{dt} = CU_B$$

Multiplication of the integral with the term $\frac{T}{T}$ with T = period time of the switching time:

$$Q = T * \left[\frac{1}{T} \int_0^T i_B \, \mathrm{dt} \right] = CU_B$$

The term in brackets is equivalent to the average supply current $I_B = \frac{1}{T} \int_0^T i_B \, \mathrm{dt}$.

$$Q = T * I_B = CU_B$$

The resulting expression for the average supply current is:

$$\underline{I_B = \frac{CU_B}{T} = \frac{U_B}{R} \text{ with } \quad R = \frac{T}{C}}$$

The result of the derivation was shown in Figure 2. Without making any proximity the equivalent resistance R is independent of the output voltage U_{Cload} and therefore even from the load itself (including C_{load} and the rectifier G) .

The input power P_R is only dependent on the intermediate voltage U_B . This dependency may be suppressed by a simple pre- filter which steers the switching frequency $f = \frac{1}{T}$.

$$P_R = \frac{U_B{}^2}{R} = \frac{CU_B{}^2}{T} \quad \text{with } R = \frac{T}{C}$$

$$T = \frac{CU_B{}^2}{P_R}$$

With a desired power consumption P_R , the capacitance of the resonance capacitor C and the measurement of the intermediate voltage U_B the pre- filter $T = \frac{CU_B{}^2}{P_R}$ will calculate the period time T of the switching frequency of the Switches S_1 , S_2 .

Without having a regulator the power consumption is independent of any kind of load and independent of the mains voltage U_N .

To test the half bridge circuit a 300 kW prototype consisting out of four 75kW power modules has been build at DESY. Single power modules have been tested successfully, but so far there was no 300kW test.

The dependence of the input power to the capacitor voltage is shown in Figure 5. The curve forms are measured values of a small power supply prototype.

2.1 Pictures of the modulator power supply

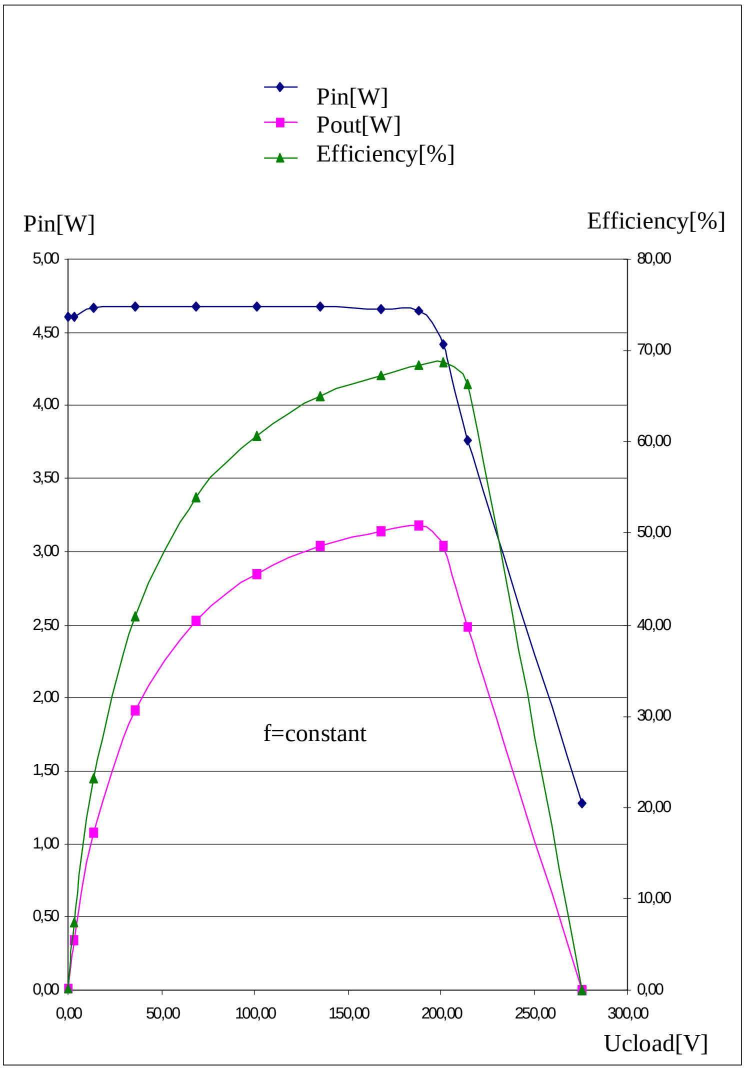


Figure 5: Power supply test circuit measurements

Conditions:

- 40W test power supply with system specific constant power consumption.
- Constant switching frequency f and varying output voltage U_{Cload} .
- No regulator or pre- filter is used.
- Because the small power prototype contains a 50Hz transformer instead of a 20kHz transformer the efficiency is 69% at maximum.

Without regulation or pre- filter the input power P_{in} is constant in a wide range of output voltages U_{Cload} = 0V to U_{Cload} = 200V. It is not possible to pass energy to a short circuit output or open circuit output. Therefore the efficiency is zero at U_{Cload} = 0V and U_{Cload} = 275V.

The test circuit proofs the formula $P_{in} = \frac{CU_B^2}{T}$ of the derivation even with 'worst case components' like the 50Hz transformer.

Figure 6: 300kW switch mode power supply including four 75 kW modules

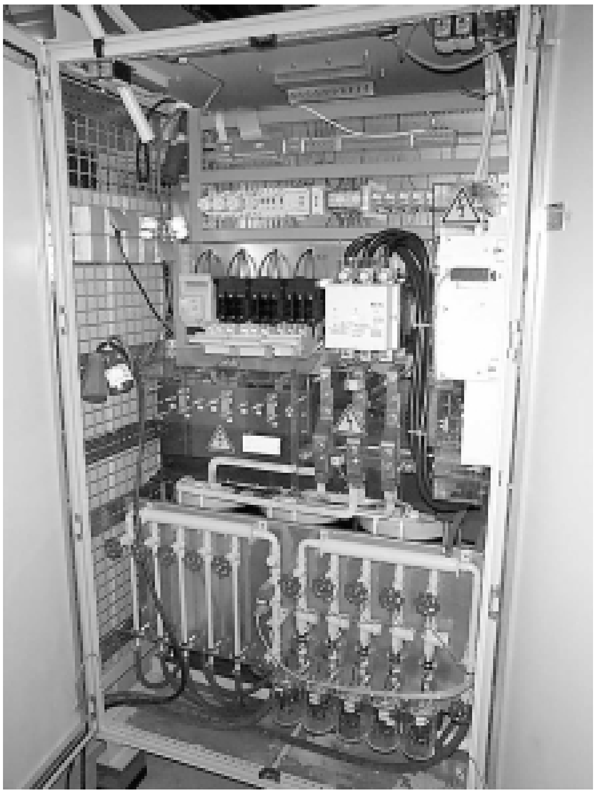
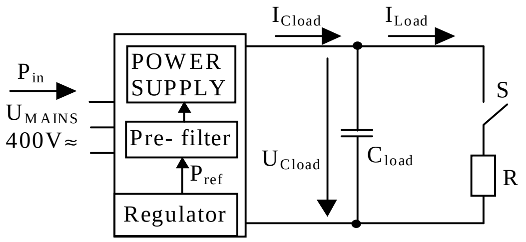


Figure 7: rear side of 300 kW switch mode power supply

3. REGULATION

The regulation is a major part of a constant power power supply. The voltage of the capacitor bank at the trigger time of the pulses has to have a pulse to pulse repetition accuracy of +/- 0.5%. This in combination with the demand of constant power requires a digital regulation. To be able to react on variations of the mains, temperature effects or non linear behavior of components the regulation is self learning. A simplified modulator circuit is shown in Figure 8.



R	Klystron
S	high voltage switch
C_{load}	main capacitor bank
U_{Cload}	capacitor voltage
P_{in}	input power

$$W_{Cload} = \frac{1}{2} C_{load} * U_{Cload}^{\,2}$$

Figure 8: Simplified modulator circuit

3.1 Principle of regulation

The regulator contains a RAM wherein the charging curve of the capacitor voltage U_{Cload} is stored. The capacitor voltage U_{Cload} is driven according to the RAM curve. A fast regulator ensures that the RAM curve and the capacitor voltage U_{Cload} are equal despite of even fast voltage variations of the mains voltage U_{MAINS} . For this reason it is obvious that the final voltage $U_e = U_{Cload}$ at time t_e remains the same at each charging cycle because it equals the well known RAM curve. Figure 9 shows the voltage curve of the main capacitor bank.

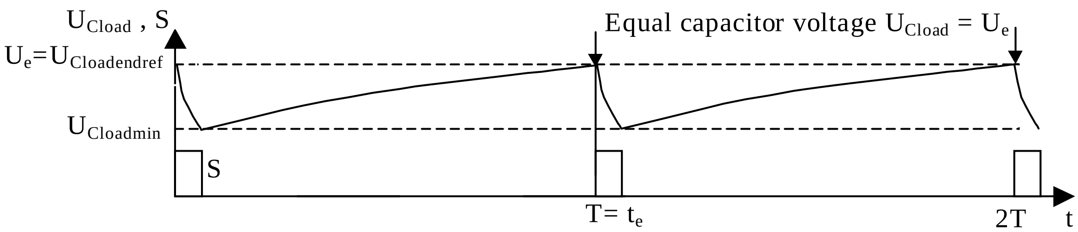


Figure 9: Voltage curve of the main capacitor

To achieve constant input power a learning process is introduced. A voltage charging curve for C_{load} is determined. With this curve the final charging voltage equals the nominal voltage (V_{capend}=$V_{capendref}$). This charging curve is stored in the memory. The possible starting curve is shown in Figure 10.

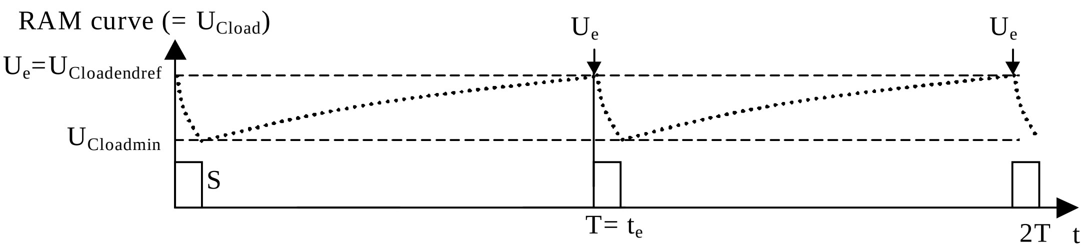


Figure 10: Stored reference charging curve of the capacitor

With this reference charging curve the input power is not yet constant but looks e.g. like the curve shown in Figure 11. The aim is to have constant power. With the self learning algorithm the stored

reference curve is modified in such a way that the reference values are increased or decreased until the input power is constant.
The result of the learning process is equal output voltage $U_{Cload} = U_e$ = constant after each charging cycle T and constant input power consumption $P_{in} = P_{average}$ = constant.

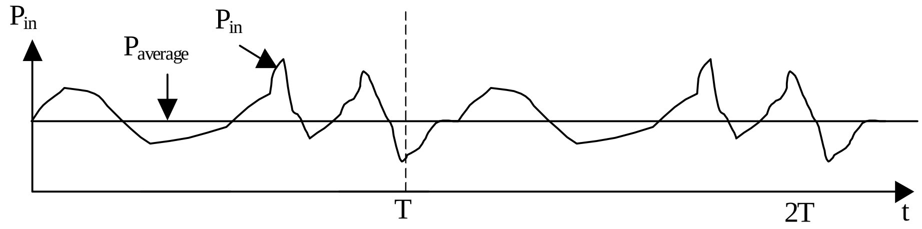


Figure 11: Input power of the modulator

3.2 Regulation with unit RAM curve

To be independent from changes in the charging time T (changes in repetition rate) or the output voltage U_e a unit curve is stored in the RAM. By this it is not necessary to relearn this curve in case of change. The unit charging curve is simply scaled to the real time and voltage axis.

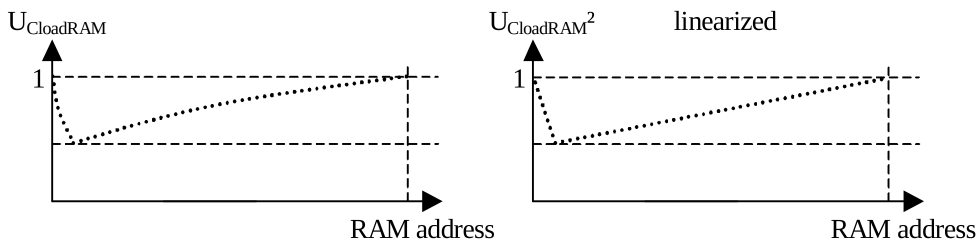


Figure 12: RAM curves of the capacitor reference curves

To linearize the regulation the capacitor voltages are used as squared values. This is because the energy stored in the capacitor C_{Load} is proportional to the squared capacitor voltage U_{Cload}^2
The linearized curve is stored as RAM curve.

$$W_{Cload} = \frac{1}{2} C_{load} * U_{Cload}^2$$

$U_{Cloadref}^2(t)$ (reference curve) is the linearized reference value of the squared capacitor voltage $U_{Cload}^2(t)$ (actual curve) .

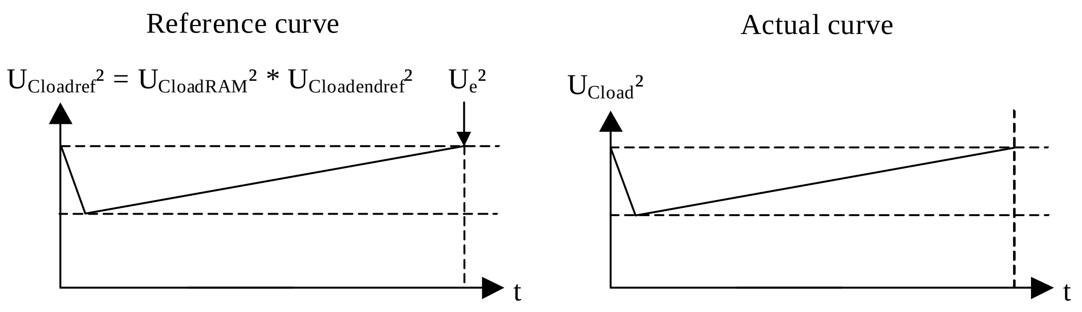


Figure 13: Linearized and scaled reference curve and monitored curve

The reference value $U_{Cloadref}^2$ and the linearized squared capacitor voltage U_{Cload}^2 are passed to a P-regulator.

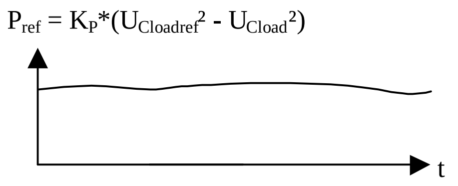


Figure 14: Output signal of the regulation

The signal P_{ref} is the signal steering the power consumption of the power supply (Figure 14). It is necessary that the actual input power P_{in} is proportional to the steering signal P_{ref} ($P_{in} \sim P_{ref}$). In some power supplies a pre- filter has to be used to ensure that $P_{in} \sim P_{ref}$. Droops of the mains voltage U_{MAINS} are the most disturbing error signals for the P(I) regulator. The pre- filter in the proposed power supply for TESLA guarantees the independence of the input power P_{in} from the mains voltage U_{MAINS} ($P_{ref} \sim P_{in} \neq f(U_{MAINS})$).
The hardware of the regulation is based on a programmable ALTERA device. The device contains the complete regulation software, the pulse firing generation of the switched mode power supply and the interlocks for the power supply. The designed board is able to work in local mode. A VME interface allows to communicate with a control computer to feed in pre- calculated curve forms for the modulator. By this the learning time is decreased.
The regulation scheme is shown in Figure 15.

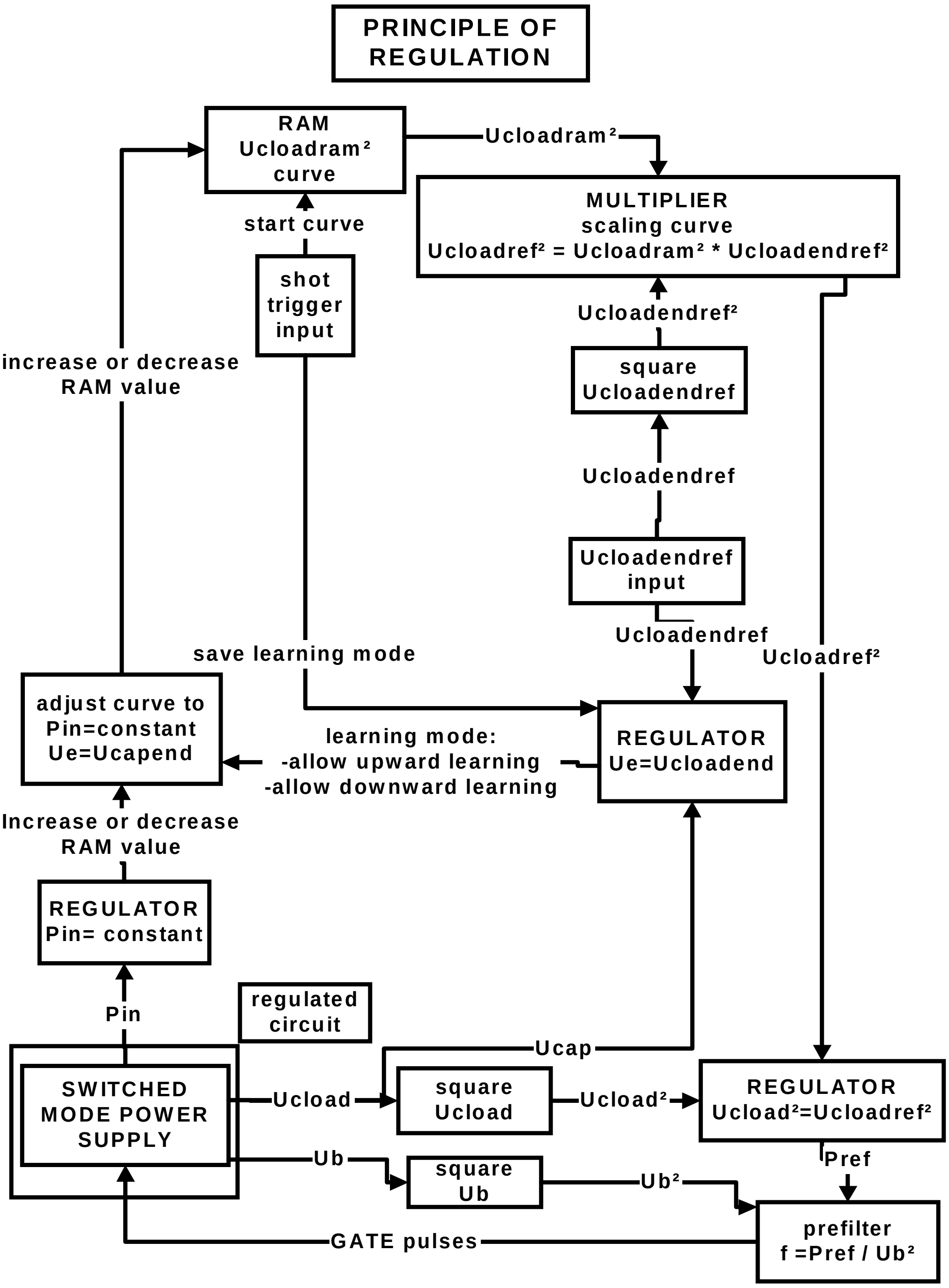


Figure 15: Block diagram of the self-learning regulation

The software modules shown in Figure 16 are implemented in the Altera FPGA device.

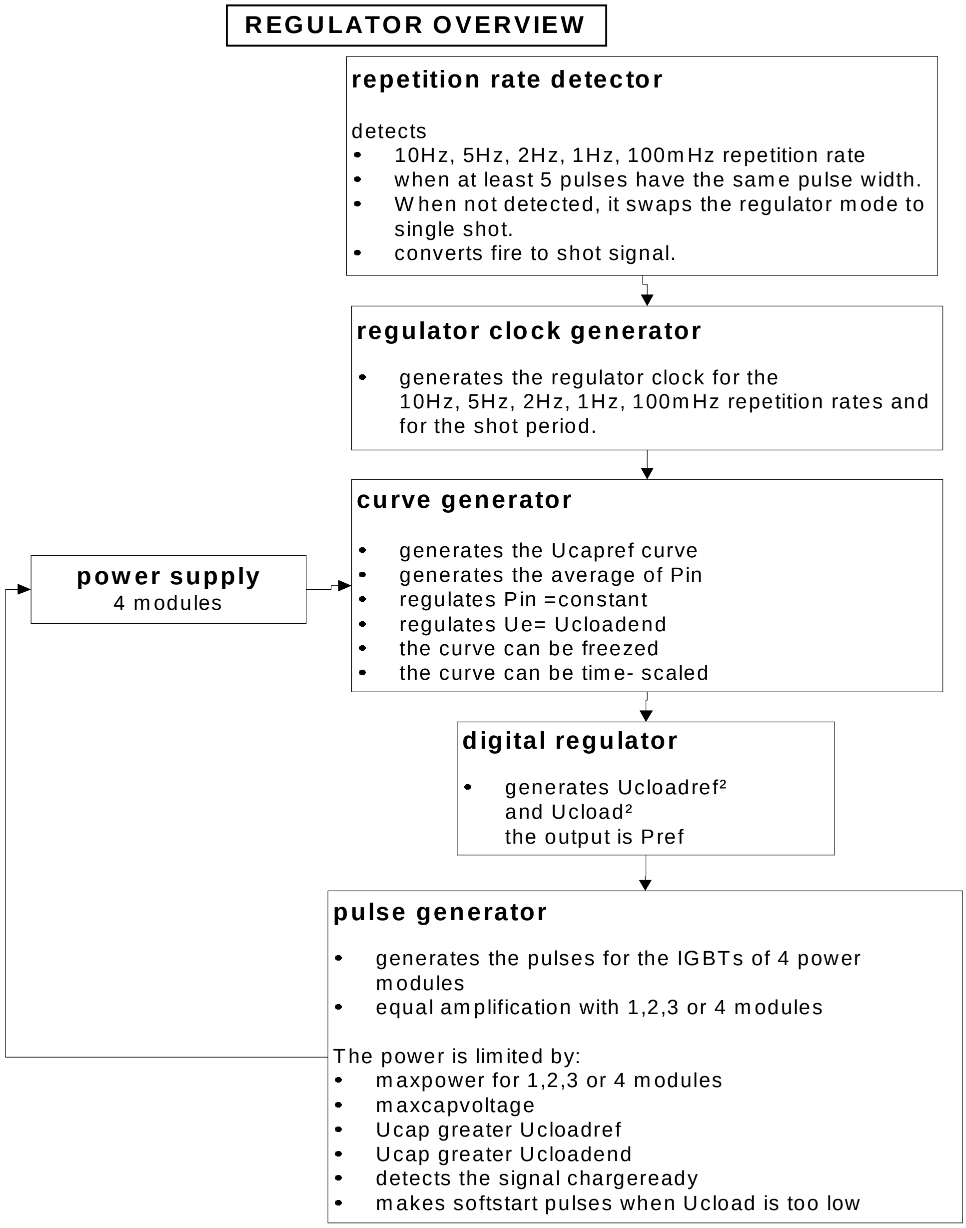


Figure 16: functional blocks in the Altera FPGA

The circuit board of the digital regulation is shown in Figure 17.

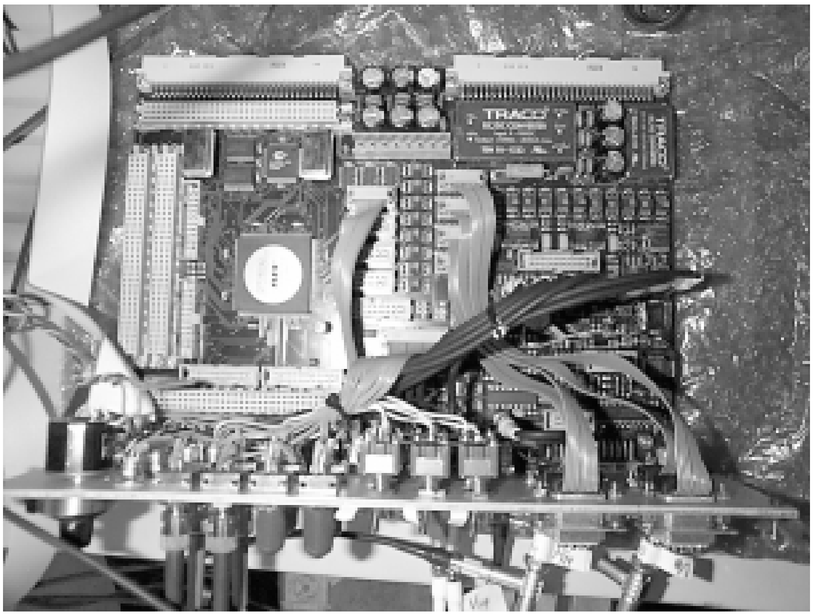


Figure 17: Top view of the digital regulator

Summary

The two different methods of power supplies have been developed and tested at DESY. The switch mode power supply already works with the desired precision using an analog regulator.
The digital regulator improves the rejection of fast disturbances such as fast droops of the mains voltage. The learning algorithm fits the charging curve to every working condition. The VME interface gives the option of downloading adjusted charging curves that are eliminating the voltage droop in the klystron voltage pulse.
Tests with other types of power supplies will follow in the future. The new self learning digital regulation will be a main part in other circuits because other power supplies may need complex pre-filters and might not be linear.

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DESIGN REQUIREMENTS FOR THE VARIOUS TYPES OF CAPACITORS USED IN MODULATORS

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Abstract

The various types of capacitor used in modulators all have very different design requirements and require differing dielectric characteristics. These capacitors range from the high voltage storage types used in the classical Pulse Forming Network modulator and the newer MOSFET modulators, the types used in the actual PFNs and the newer requirements of relatively low voltage types and IGBT snubbers which are both used in the CLW modulator. These design requirements are discussed in conjunction with the advantages and disadvantages of the different types of modulator.

INTRODUCTION

The various types of capacitor used in modulators all have very different design requirements and require differing dielectric characteristics. As the capacitors have a critical role to play in the efficient and reliable working of the modulator, attention to their design is of great importance. Often circuit designers regard capacitors as necessary evil devices that take up a large amount of the total volume available in the modulator. They are therefore tempted to use capacitors that were not designed for the job, risking long-term unreliability for the whole modulator. Reliability and long life are paramount and should not be traded for size reduction or cheapness.

Some types of modulator require less sophisticated and others more complex capacitors. The varying design requirements are discussed in conjunction with the advantages and disadvantages of the different types of modulator.

The capacitor parameters that are important are: -

Reliability and life

Capacitance tolerance

Capacitance temperature coefficient

Capacitance voltage coefficient

Inductance of the capacitor

Losses

- Effective Series Resistance
- Dielectric loss and its frequency dependence

Size

- Energy density

Each of these parameters assumes a different degree of importance depending on the circuit requirements and some are mutually exclusive. We therefore need to know in detail the circuit parameters such as: -

Is the application basically DC or are there AC components?

If there are AC components, what is the waveform? In particular, one must consider:

The repetition frequency.

The rate of change of voltage and current

The choice of a suitable dielectric is necessary to fulfil the capacitor parameter requirements listed above. If tolerance is an important parameter we would then not choose a ceramic dielectric as it is very difficult to manufacture such capacitors to close accuracy and they usually have both a high temperature and voltage coefficient. However, ceramic capacitors usually have low dielectric losses (plus the capability of withstanding high temperatures) and are therefore often suitable for RF use. Similarly, polyester film has relatively high temperature capability but its dielectric loss is frequency-dependent. This dielectric is therefore suitable for capacitors that are basically DC. Polypropylene has a lower temperature capability but its loss angle is low and nearly independent of frequency. This dielectric is therefore used in AC and many pulse applications. Mica has great stability, is capable of withstanding very high temperatures and has a very low loss angle over a large frequency range. However, capacitors constructed using mica as a dielectric have an extremely low energy density and, as a result, are large and very expensive.

1. CAPACITORS FOR CLASSICAL PULSE FORMING NETWORK MODULATORS

In a classical PFN modulator capacitors can be used as a reservoir in the charging circuit and are used in the PFN for pulse shaping. The design of the two types of capacitor is totally different, as, in the first case, the capacitor is basically a high voltage DC device, whereas in the second they are subjected to pulse operation.

For the reservoir capacitor the dielectric normally used is a mixture of oil-impregnated paper and polyester film. The paper is used between layers of polyester and between the aluminium foil and the polyester to act as a wick allowing oil impregnation of the capacitor element. However, as the insulation resistance of the polyester is several orders higher than that of the oil impregnated paper, nearly all the DC stress falls on the polyester and very little on the paper and the interfaces between the various layers; this means that damaging ionisation is avoided. A typical reservoir capacitor working at, say, 20kV would have four series sections each working at 5kV with the stress on the polyester film approaching 200V per micron. The energy density of such a unit would be around 68 Joules per litre if lifetime is to be several tens of thousand hours. Of course, not all the total volume is taken up by the dielectric, as major insulation to the metal case to be able to hold off 20kV is also required. Furthermore, clearance must be allowed for the HT terminal and this adds to the overall space requirement.

However, the PFN capacitors are totally different as in normal operation; at each pulse discharge, they are subjected to a rapid change of voltage from their working level to around zero or just below zero. The pulse current resulting from these voltage changes is high, as are the component frequencies. It is therefore important to use a dielectric that has a low loss angle over a wide frequency spectrum. Furthermore, the rapid change of voltage means that damaging ionisation within the dielectric layers is a constant problem and it has been found over the years that this can be avoided if the number of capacitor elements in series is high and the energy density low. Typically, a PFN capacitor working at around 20kV would have 20 capacitor elements in series. Each element would use polypropylene and oil impregnated paper as the dielectric; polypropylene is chosen as it has extremely low dielectric losses over a wide frequency range and the losses, in the capacitive part of the PFN, are thus kept to a minimum. However, the trade-off for this is an increase in volume and a very significant reduction in energy density, to a figure as low as 3 Joules per litre for lifetimes of several tens of thousand hours.

This classical PFN modulator, apart from being much larger in volume for a given output power than some of the newer types, has the disadvantage of producing a fixed pulse length that cannot be altered without considerable difficulty. A further complication is the use of a thyratron as the switching device. This component has a relatively short life measured in a few thousands of hours and this can be substantially shortened by damage caused by load arcs.

2. CAPACITORS USED IN SERIES SOLID STATE SWITCHED MODULATORS

This title covers several different attempts to bring solid state technology to modulator design with varying degrees of success.

In its simplest form the thyratron in the classical PFN modulator, discussed earlier, is replaced by number of solid state switches connected in series. These switches could be either thyristors or IGBTs. Most attempts have proved that this technique is unreliable and this is mainly because of the problems associated with ensuring that all the switches fire at the same moment and the limitations of dI/dt associated with thyristor turn-on physics. If one switch is slower than all the others, for whatever reason, (or is not triggered at all due to trigger system failure) all the stress will be developed across that one switch which will then, inevitably fail. The problems worsen as the voltage of the required from the modulator is increased and the number of series semiconductor elements increases in proportion.

A technique that may prove to be successful for low power modulators is one that uses MOSFETS in series. These MOSFETS are run well below their rated voltage and are laddered with a network of resistors and capacitors to ensure voltage sharing even if there is some jitter in the switching. The pulse is supplied by a high voltage capacitor, charged to the pulse voltage, and which is then discharged into the load - often a magnetron – with a pulse shaping network on the output. The capacitor used in this type of circuit will typically be required to store between ten and twenty times the energy required in each pulse so that it will only "droop" a relatively small amount from its working voltage during the discharge pulse. It will have to be a relatively low inductance capacitor or, rather the capacitor with its discharge switch bank will have to have relatively low inductance, in order to achieve a fast rising pulse. The design of the storage capacitor for this type of modulator is somewhat problematical as it has to have the high voltage major insulation required to protect the capacitor from flashing over to its case or some other nearby low potential point. The problem arises because space is usually at a premium. This may be overcome by submerging the whole modulator in an oil tank thus allowing clearances to be reduced. But this in itself produces a further problem as having the whole modulator under oil normally means that on-site servicing is not possible. As the whole unit is composed of many hundreds of discrete components failure of some component will be likely to occur and, if the failed part is in a critical position, this will necessitate the whole modulator being removed from service and returned to the manufacturer for rectification.

The capacitor used for this duty would normally be very similar to one used for DC duty but with high pulse current capability and low Equivalent Series Resistance (ESR). The low ESR is needed to keep the losses in the capacitor low when it is being discharged. The high current capability is normally obtained by using multi-tabs or extended foil. Care has to be taken when the modulator is working at high repetition rates to ensure that the capacitor losses (normally series resistance losses) do not cause the whole modulator assembly to overheat. The dielectric would normally be mixed polyester and paper as discussed earlier for the Classical PFN modulators. The stress in the polyester film could be around 200V per micron but the energy density would fall to around 50 joules per litre due to the extra requirement for major insulation. Of course, this would be effectively reduced much further as the capacitor is discharged only by between 5 and 10% of its working voltage at each pulse.

3 CAPACITORS USED IN THE NLC AND CLW MODULATORS.

Although these modulators are rather different they both use very similar capacitors. We have therefore combined the discussion of the capacitors required for both these modulators but in order to clarify the requirements, we will first discuss the basics of these two new-concept circuits.

In the NLC modulator the secondary of the pulse transformer is composed of a single tube and is, in effect, a single turn. A very suitable voltage for the current range of IGBT switches is around 1kV. If we assume the output voltage is 200kV we would need a voltage magnification of 200:1 so that the primary will have to be 1/200 turns. One two hundredth of a turn is obviously impractical but this can effectively be achieved by having two hundred individual primary coils each with its own magnetic core. These primaries are all discharged at the same time and the resulting high voltage and high power pulse is used to drive the load tube or indeed tubes. This type of modulator is extremely good for very high power pulses provided they are of short duration. The problem arises when relatively long duration pulses are needed, as the amount of magnetic material required becomes extremely high.

In the CLW modulator the fractional primary-turn concept is achieved by winding many individual primaries on several sections of one core whilst the multi-turn secondary is wound around the whole core. Full details of this modulator will be discussed in the paper to be presented here tomorrow by my colleague Walter Crewson. This modulator can be designed to give long pulses and average power up to around 1MW.

In both these modulators capacitors rated at between 1 and 2kV are charged and then discharged using IGBTs to switch the capacitor current both on and off. These capacitors are discharged into the multi-primaries of the pulse transformer with the load tube connected to the single secondary. As in the previous modulator the capacitors are only discharged by around 5% to 10% of their total voltage but because the total voltage is only up to 2kV, metallised polypropylene is found to be a suitable dielectric. This system has the great advantage that it is self-healing and can therefore be used at very high stress and energy density. Work has been started in the use of these capacitors for producing the fast front edge of the pulse and then switching to a large reservoir bank of electrolytic capacitors for sustaining the pulse when long pulses are required. This technique could substantially reduce the overall cost and size of a long pulse modulator but further research needs to be done. Of course, in both the NLC and the CLW modulator the magnetic core volume to sustain long pulses is increased substantially if saturation is to be avoided.

IGBTs used for this type of duty are working at a fraction of their average power rating but are considered safe to be used at twice their maximum current rating when used for pulse duty provided that their voltage rating is not exceeded. Problems arise when a fault in the load such as arcing occurs. IGBTs have a turn-off delay and during this delay time the large current rise through the IGBT can be damaging and in order to reduce the turn-off (LdI/dt) voltage spike it is essential to fit a snubber capacitor between the emitter and the collector of the IGBT. These capacitors must have very low inductance if they are to work effectively and the diode connected in series with the snubber has to have a very fast turn-on if it is start conducting before damage occurs. Effective snubbers have been designed using metallised polypropylene as the dielectric with strip copper lead-outs to reduce the inductance.

This paper has discussed in general terms the capacitor design requirements for a number of different types of modulator and illustrates the reasons why different dielectrics and techniques are used.

FIRST RESULTS OF THE LOS ALAMOS POLYPHASE BOOST CONVERTER-MODULATOR

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Abstract
This paper describes the first full-scale electrical test results of the Los Alamos polyphase boost converter-modulator being developed for the Spallation Neutron Source (SNS) at Oak Ridge National Laboratory. The converter-modulator provides 140 kV, 1.2 ms, 60 Hz pulses to a 5 MW, 805 MHz klystron. The system, which has 1 MW average power, derives its +/- 1250 Volt DC buss link voltages from a standard 3-phase utility 13.8 kV to 2100 volt transformer. An SCR pre-regulator provides a soft-start function in addition to correction of line and load variations, from no-load to full-load. Energy storage is provided by low inductance self-clearing metallized hazy polypropylene traction capacitors. Each of the 3-phase H-bridge Insulated Gate Bipolar Transistor (IGBT) Pulse-Width Modulation (PWM) drivers are resonated with the amorphous nanocrystalline boost transformer and associated peaking circuits to provide zero-voltage-switching characteristics for the IGBT's. This design feature minimizes IGBT switching losses. By PWM of individual IGBT conduction angles, output pulse regulation with adaptive feedforward and feedback techniques is used to improve the klystron voltage pulse shape. In addition to the first operational results, this paper will discuss the relevant design techniques associated with the boost converter-modulator topology.

1. INTRODUCTION

The simplified bock diagram of the converter/modulator system is shown in Figure 1. This system minimizes costs with the utilization of many standard and proven industrial and utility components. The substation is a standard 3 phase 13.8 kV to 2100 V vacuum-cast core transformer with passive harmonic traps and input harmonic chokes. These components are located in an outdoor rated NEMA 3R enclosure that does not require secondary oil containment or related fire suppression equipment. The Los Alamos prototype is manufactured by *Dynapower Corporation* in Burlington, Vermont. The power transformer is followed by an SCR pre-regulator that accommodates incoming line voltage variations and other voltage changes resulting from transformer and trap impedances, from no-load to full-load. The SCR pre-regulator also provides the soft-start function. The SCR regulator provides a nominal +/- 1250 Volt output to the energy storage capacitor banks. The SCR pre-regulator utilized in the Los Alamos prototype is manufactured by *NWL* in Bordentown, New Jersey. The energy storage capacitors are self-clearing metallized hazy polypropylene traction motor capacitors. As in traction application, these capacitors are hard bussed in parallel. These capacitors do not fail short, but fuse or "clear" any internal anomaly. At our capacitor voltage rating (1.5 kV) there has not been a recorded internal capacitor buss failure. In this application, as in traction motor applications, the capacitor lifetime is calculated to be 1e9 hours, before de-rating factors are included. A special low inductance design for these capacitors has been developed by *Thomson Components* in Saint-Apollinaire, France. The Insulated Gate Bipolar Transistors (IGBT's) are configured into three "H" bridge circuits to generate a three phase 20 kHz square wave voltage drive waveform applied to the transformer primaries. The IGBT's are "chirped" the appropriate

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duration to generate the high voltage klystron pulse, typically 1.2 ms. Due to the IGBT switching power levels, currents, and frequencies involved, low inductance busswork and bypassing is of paramount importance. The IGBT's are 3300 volt, 1200 amp devices (*FZ1200R33KL2)* manufactured by *EUPEC* of Hanau, Germany. The boost transformers utilize amorphous nanocrystalline material that has very low loss at the applied frequency and flux swing. Operating at 20 kHz and about 1.6 Tesla bi-directional, the core loss is about 1.2 watts per pound in our application, or 320 W per core. Each of the "C" cores (one for each phase) weigh 260 lbs. and has a 3.5" by 5" post. The nanocrystalline material is manufactured by *AMET,* located in Asah, Russia. By appropriately spacing the secondary from the primary, the transformer leakage inductance can be resonated with secondary shunt peaking capacitors to maximize voltage output and tune the IGBT switch current to provide "zero-voltage-switching" with IGBT turn-on. The zero-voltage-switching occurs when the IGBT gate drive is positive, but reverse transformer primary circulating current is being carried by the IGBT internal freewheel diode. We have tuned for about 4 μs of freewheel current before the IGBT conducts in the normal quadrant. This tuning provides for about 15% control range (4/25 μs) for IGBT pulse width modulation (PWM). Further transformer design optimizations can change IGBT commutation (turn-off) current for control range and coupling coefficient for IGBT peak current. As transformer design characteristics interact with other circuit parameters, optimization may be performed for various klystron loads and voltages. IGBT PWM of the active klystron voltage pulse enables us to use adaptive feedback and feedforward techniques with digital signal processors (DSP's) to regulate and provide "flat" klystron voltage pulses, irrespective of capacitor bank "start" voltage and related droop. The DSP adaptive feedback/feedforward processor used in the Los Alamos prototype was manufactured by *Z-TEC Inc.*, of Albuquerque, New Mexico. Line synchronization is not absolutely required as the adaptive DSP can read bank voltage parameters at the start of each pulse and calculate expected droop. The output high-voltage rectification circuit is a standard six-pulse rectification circuit with a "pi-R" type filter network. The diodes are high-voltage fast recovery ion-implanted types, manufactured by *IXYS*, which are series connected with the appropriate compensation networks. The diodes have the second highest total power loss (after the IGBT's) and are forced oil cooled. The filter network must attenuate the 120 kHz switching ripple and have a minimal stored energy. The stored energy is wasted energy that must be dissipated by the klystron at the end of each pulse. With the parameters we have chosen, the ripple is very low (~300 volts) and the klystron fault energy (in an arc-down) is about 10 joules. Even if the IGBT's are not turned off, the transformer resonance is out of tune in a fault condition, and little difference in klystron fault energy will result. If the IGBT's fail short, through the transformer primary winding, the boost transformer will saturate in about 30 uS, also limiting any destructive faults to the klystron. In a faulted condition, the klystron peak fault current is about twice nominal, with low dI/dT's.

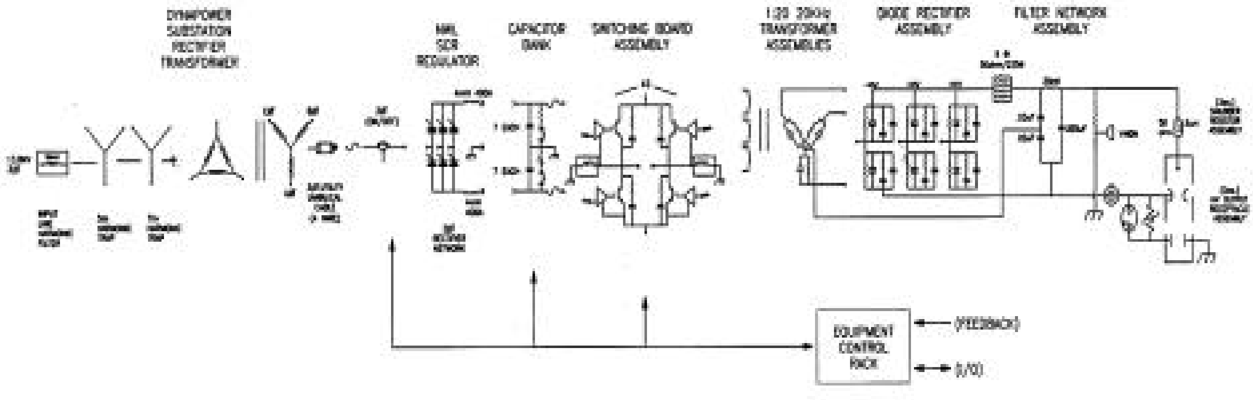


Figure 1 Simplified Block Diagram

2. MODELING

The complete electrical system of the converter/modulator system has been modeled in extreme detail. This includes design studies of the utility characteristics, transformer and rectification methodology (e.g. 6 pulse vs. 12 pulse), IGBT switching losses, boost transformer parameters, failure modes, fault analysis, and system efficiencies. Various codes such as SCEPTRE, MicrocapIV, Flux2D, and Flux3D have been used to perform these tasks. SCEPTRE has been primarily used to examine IGBT and boost transformer performance in great detail to understand design parameters such as switching losses, IGBT commutation dI/dT, buss inductance, buss transients, core flux, core flux offset, and transformer Eigen frequencies. Flux2D and Flux3D have been used to examine transformer coupling coefficients, leakage inductance, core internal and external flux lines, winding electric field potentials, and winding field stresses. The Flux2D and Flux3D were particularly useful to examine transformer secondary winding profiles that gave the desired coupling coefficients with minimized electrical field stresses. Micro-CapIV has been used to examine overall design performance of the system. This includes the utility grid parameters such as power factor, line harmonics, and flicker. We have optimized the design to accommodate the IEEE-519 and IEEE-141 harmonic content and flicker standards. Micro-CapIV uses simplified switch models for the IGBT's, which does not accurately predict their losses. However, the code has been very useful to examine tradeoffs of circuit performance with the lumped elements such as the boost transformer, shunt peaking capacitance, the filter networks, and the input energy stores. Micro-CapIV is also adept at making parametric scans to determine component sensitivities and tolerances. Comparisons between the SCEPTRE and MicrocapIV codes show no significant differences in the system operational performance such as switching currents, switching voltages, and output voltage.

3. FIRST RESULTS

The converter-modulator made its first high voltage pulse on January 17, 2001. After 10 days of testing and replacement of defective dummy load components, full pulse output voltage (140 kV), pulse width (~1.2 ms), and peak power (11 MW) were obtained. As shown in Figure 2, the 140 kV risetime is about 20 μs with about a 6% bank droop. The risetime compares favorably to other long-pulse modulator topologies. Additional high-voltage tests were made at 80 kV, as this voltage will be used in the SNS linac superconducting portion. The 80 kV operations are shown in Figure 3. 80 kV operations with the DSP adaptive feedback/feedforward processor is shown in Figure 4. The DSP processor removes all overshoot and bank voltage droop.

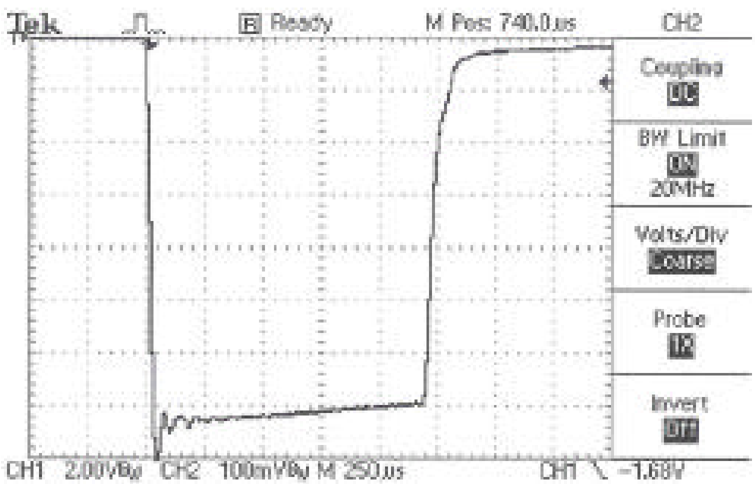


Figure 2 140 kV Output Pulse, 20 kV/Division

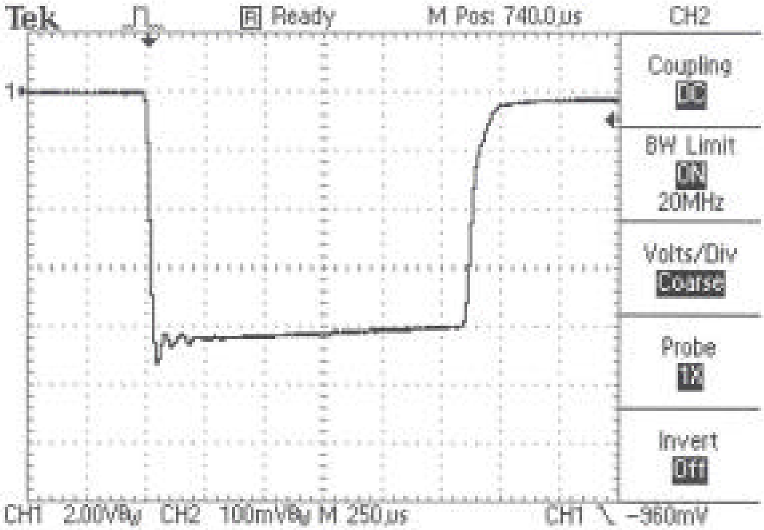


Figure 3 80 kV Output Pulse, 20 kV/Division

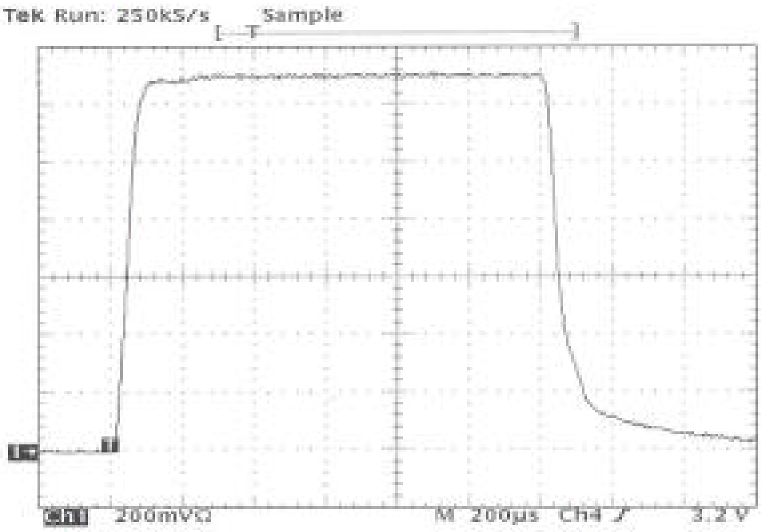


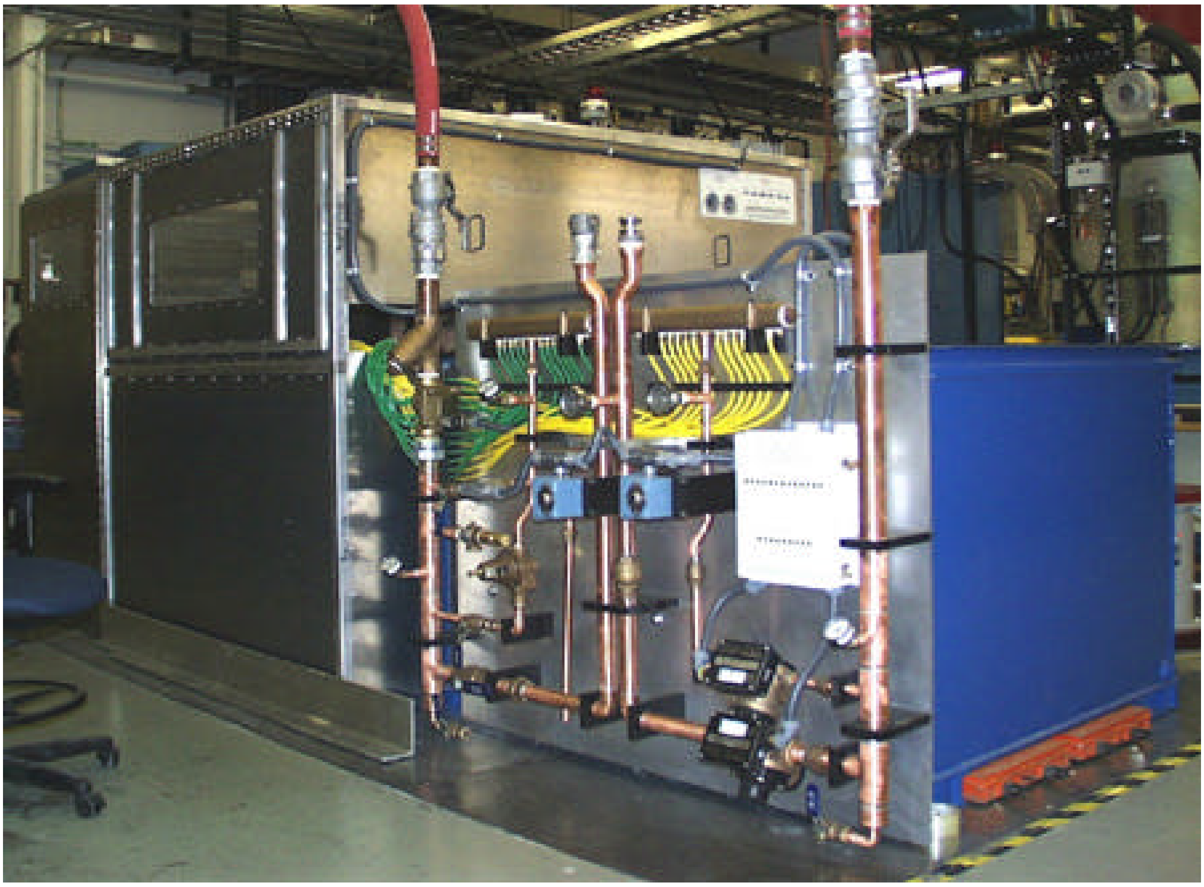
Figure 4 80 kV Output with Adaptive Feedback/Feedforward

2. PROJECT STATUS

The low average power testing of the first article has been completely successful. The substation construction work is nearing completion and it is anticipated that high average power testing will begin in May.

3. CONCLUSION

The converter/modulator has demonstrated several new design methodologies that are expected to revolutionize long-pulse klystron modulator design. These items include special low-inductance self-clearing capacitors, large amorphous nanocrystalline cut-core transformers, high-voltage and high-power polyphase quasi-resonant DC-DC conversion, and adaptive power supply control techniques. The first test results on the initial design were achieved in about a year after conception. Design economies are achieved by the use of industrial traction motor components (IGBT's and self-clearing capacitors) and standard utility cast-core power transformers. The compact and modular design, Figure 5, minimizes on-site construction and a simplified utility interconnection scheme further reduces installation costs. The design does not require capacitor rooms and related crowbars. By generating high-voltage only when needed, reliability and personnel safety is greatly enhanced. This approach provides design flexibility to operate klystrons of different voltages primarily by changing the boost transformer turns ratio. Other optimizations also permit "CW" operation of the polyphase boost converter topology. All testing of the full-scale system have been completely successful and all results agree with modeling efforts to date, which indicate the design methodologies will be imminently successful.



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