CAS
CERN ACCELERATOR SCHOOL
POWER CONVERTERS FOR PARTICLE ACCELERATORS

Hyatt Conference Centre, Montreux, Switzerland
26–30 March 1990

PROCEEDINGS
Editor: S. Turaci

GENEVA
1990
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ABSTRACT

This volume presents the proceedings of the fifth specialized course organized by the CERN Accelerator School, the subject on this occasion being power converters for particle accelerators. The course started with lectures on the classification and topologies of converters and on the guidelines for achieving high performance. It then went on to cover the more detailed aspects of feedback theory, simulation, measurements, components, remote control, fault diagnosis and equipment protection as well as systems and grid-related problems. The important topics of converter specification, procurement contract management and the likely future developments in semiconductor components were also covered. Although the course was principally directed towards DC and slow-pulsed supplies, lectures were added on fast converters and resonant excitation. Finally the programme was rounded off with three seminars on the related fields of Tokamak converters, battery energy storage for electric vehicles, and the control of shaft generators in ships.
CERN ACCELERATOR SCHOOL

will organize a course on

POWER CONVERTERS FOR
PARTICLE ACCELERATORS

26–30 March 1990
at the Hyatt Conference Centre, Montreux, Switzerland

Lectures:
Performance requirements for accelerators
Classification of converters
DC and slow-pulsed converter topologies
Fast-pulsed power converters
Achieving high performance
Feedback theory
Transducers and measurement technology
Simulation
Switch-mode
Wound components
Power devices and drivers
System aspects
Remote control and fault diagnosis
Protection interlocks
Grid-related and interference problems
Resonant excitation
Procurement
Development of new ideas

Seminars:
Tokamak plasma control using thyristor power converters
Simulation model for thyristor-controlled shaft generators
Battery energy storage for electric vehicles

Visit:
CERN, European Organization for Nuclear Research

General Information
This course will mainly be of interest to staff in accelerator laboratories, university departments, and manufacturing companies specializing in power converters and their electronics. It will concentrate on d.c. and slow-pulsed converters via lectures, seminars, and a visit to CERN. Persons wishing to attend this School can obtain further information and application forms from:
Ms. S. von Warburg, CERN Accelerator School, LEP Division, CH-1211 Geneva 23,

or by electronic mail: CASPOWER@CERNVM.CERN.CH

Application forms must be returned no later than 15 December 1989. The registration fee is CHF 1300 per person if sharing a double room, and includes full board and lodging. For rates for single rooms, local students, and for non-participating accompanying persons (space permitting), please see the application form. All participants will receive a copy of the proceedings of this course.

Administration

Sponsored by: HOLEC HH projects P.O. Box 258, Hengelo, Holland. Tel: 31.74.469111
# POWER CONVERTERS FOR PARTICLE ACCELERATORS

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* Paper by H. Bühler
FOREWORD

Particle accelerators touch many disciplines and technologies and the CERN Accelerator School (CAS) tries to reflect this diversity through its specialized courses. The most recent of these was held in Montreux, 26–30 March 1990, and focused on DC and slow-pulsed power converters for particle accelerators.

Accelerators and power converters have always been closely associated. In fact the first accelerator capable of nuclear physics research, built by Cockcroft and Walton, was mainly an exercise in building a high-voltage power converter. Such was the success of their solution that the Cockcroft and Walton Generator became the preferred pre-injector for large accelerators for many decades after. Today, power converters have diversified into all the subsystems of large accelerators and proliferated, until in a machine like LEP there are literally hundreds of independent converters powering the guide field alone. The consequent demands made on reliability and performance have been a major driving force in the development of these devices.

The course presented in these proceedings is the result of a serious effort by the very many people involved. The continued support of the CERN Directorate for the CAS courses and the guidance of the CAS Advisory and Programme Committees are both essential and much appreciated. The Local Organizing Committee ensured that the course ran smoothly, and many of the CERN services contributed in one way or another to its success and the publication of these proceedings. Very special thanks are due to the lecturers, for the tremendous effort they put into preparing, presenting and writing up their topics. The sponsorship of Holc Innovation and Technology, Holland, was most appreciated as was the efficiency of the Hyatt's Conference Centre staff. Finally, the greatest encouragement to all concerned came from the fact that some 80 students attended the course, representing the accelerator laboratories, universities and industrial firms of no less than 16 different countries.

P.J. Bryant, Head of CAS
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CERN ACCELERATOR SCHOOL
POWER CONVERTERS FOR PARTICLE ACCELERATORS

Hyatt Conference Centre, Montreux, Switzerland
26–30 March 1990

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Editor: S. Turaci

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ABSTRACT

This volume presents the proceedings of the fifth specialized course organized by the CERN Accelerator School, the subject on this occasion being power converters for particle accelerators. The course started with lectures on the classification and topologies of converters and on the guidelines for achieving high performance. It then went on to cover the more detailed aspects of feedback theory, simulation, measurements, components, remote control, fault diagnosis and equipment protection as well as systems and grid-related problems. The important topics of converter specification, procurement contract management and the likely future developments in semiconductor components were also covered. Although the course was principally directed towards DC and slow-pulsed supplies, lectures were added on fast converters and resonant excitation. Finally the programme was rounded off with three seminars on the related fields of Tokamak converters, battery energy storage for electric vehicles, and the control of shaft generators in ships.
CERN ACCELERATOR SCHOOL

will organize a course on

POWER CONVERTERS FOR
PARTICLE ACCELERATORS

26–30 March 1990
at the Hyatt Conference Centre, Montreux, Switzerland

Lectures:
Performance requirements for accelerators
Classification of converters
DC and slow-pulsed converter topologies
Fast-pulsed power converters
Achieving high performance
Feedback theory
Transducers and measurement technology
Simulation
Switch-mode
Wound components
Power devices and drivers
System aspects
Remote control and fault diagnosis
Protection interlocks
Grid-related and interference problems
Resonant excitation
Procurement
Development of new ideas

Seminars:
Tokamak plasma control using thyristor power converters
Simulation model for thyristor-controlled shaft generators
Battery energy storage for electric vehicles

Visit:
CERN, European Organization
for Nuclear Research

General Information
This course will mainly be of interest to staff in accelerator laboratories, university departments, and manufacturing companies specializing in power converters and their electronics. It will concentrate on d.c. and slow-pulsed converters via lectures, seminars, and a visit to CERN. Persons wishing to attend this School can obtain further information and application forms from:
Msr. S. von Warburg, CERN Accelerator School, LEP Division, CH-1211 Geneva 23,
or by electronic mail: CASPOWER@CERNVM.CERN.CH
Application forms must be returned no later than 15 December 1989. The registration fee is CHF 1300 per person if sharing a double room, and includes full board and lodging. For rates for single rooms, local students, and for non-participating accompanying persons (space permitting), please see the application form. All participants will receive a copy of the proceedings of this course.

Administration

Sponsored by: HOLEC projects
P.O. Box 258, Hengelo, Holland. Tel: 31.74.469111
# POWER CONVERTERS FOR PARTICLE ACCELERATORS

Hyatt Continental, Montreux, 26-30 March 1990

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* Paper by H. Bühler
FOREWORD

Particle accelerators touch many disciplines and technologies and the CERN Accelerator School (CAS) tries to reflect this diversity through its specialized courses. The most recent of these was held in Montreux, 26–30 March 1990, and focused on DC and slow-pulsed power converters for particle accelerators.

Accelerators and power converters have always been closely associated. In fact the first accelerator capable of nuclear physics research, built by Cockcroft and Walton, was mainly an exercise in building a high-voltage power converter. Such was the success of their solution that the Cockcroft and Walton Generator became the preferred pre-injector for large accelerators for many decades after. Today, power converters have diversified into all the subsystems of large accelerators and proliferated, until in a machine like LEP there are literally hundreds of independent converters powering the guide field alone. The consequent demands made on reliability and performance have been a major driving force in the development of these devices.

The course presented in these proceedings is the result of a serious effort by the very many people involved. The continued support of the CERN Directorate for the CAS courses and the guidance of the CAS Advisory and Programme Committees are both essential and much appreciated. The Local Organizing Committee ensured that the course ran smoothly, and many of the CERN services contributed in one way or another to its success and the publication of these proceedings. Very special thanks are due to the lecturers, for the tremendous effort they put into preparing, presenting and writing up their topics. The sponsorship of Holec Innovation and Technology, Holland, was most appreciated as was the efficiency of the Hyatt's Conference Centre staff. Finally, the greatest encouragement to all concerned came from the fact that some 80 students attended the course, representing the accelerator laboratories, universities and industrial firms of no less than 16 different countries.

P.J. Bryant, Head of CAS
S. Turner, Editor
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PERFORMANCE REQUIREMENTS FOR ACCELERATORS
P.J. Bryant
CERN, Geneva, Switzerland

Abstract
The power converter played a central role in the early DC accelerators and the well-known principle of the transformer was the inspiration for an acceleration mechanism. As accelerators developed towards higher energies, in response to the needs of physics research, power converters proliferated into all the sub-systems so that literally hundreds of units will be found in a modern accelerator such as the CERN LEP. These large numbers and the critical roles the power converters play imply very high reliability and the accelerator beam optics frequently impose very severe specifications on performance. Thus accelerators have become an important driving force for the development of high-performance power converters. The relationships between accelerator performance and power converter specifications are discussed with some examples of the criteria used to set tolerances.

1. THE EARLY HISTORY OF ACCELERATORS

At the turn of the century, research into the structure of matter was advancing rapidly and particle accelerators were a natural development arising out of that research. It was Rutherford who realised the need for an accelerator and who encouraged his colleagues Cockcroft and Walton to build the first machine for nuclear physics research.

Ernest Rutherford was born in New Zealand on August 30th 1871. He went to Trinity College Cambridge in 1894 to work with the British physicist Sir Joseph Thompson, the discoverer of the electron. In 1898 Rutherford became the Professor of Physics at McGill University in Montreal, Canada. During this time he studied radioactivity and in 1906 he started experiments on bombarding mica sheets with alpha particles of several MeV from natural sources. His observations started him thinking about atomic structure and scattering. In 1907 he returned to England to become Professor of Physics at Manchester University and shortly after in 1908 he received the Nobel Prize in chemistry for his work with radioactive substances. He published his theory on atomic structure in 1911.

In 1919 he became Professor of Experimental Physics at the Cavendish Laboratory Cambridge where he directed his studies to the nucleus of the atom and induced nuclear reactions with natural alpha particles.

Natural radioactive sources are limited in energy and intensity and Rutherford realised that he needed an 'artificial' source of several MeV with which to continue his studies. At that time, the electrostatic machines of the Wimshurst type were well known, but the 100 kV attainable by such a machine fell far short of the few MV level that Rutherford believed was necessary to split the nucleus. For a few years there was no progress and then the situation changed suddenly in 1928, when Gurney and Gamov independently predicted tunnelling [1]. With tunnelling, it appeared that an energy of just 500 keV might suffice. This seemed technologically feasible to Rutherford and he immediately encouraged J.D. Cockcroft and E.O. Walton to start designing a 500 kV particle accelerator. Four years later in 1932, they split the lithium atom with 400 keV protons, which was the first fully man-controlled splitting of the atom [2]. This feat earned them the Nobel Prize in 1951 and the recognition of having built the first particle accelerator capable of nuclear research.
Fig. 1 Cockcroft and Walton's apparatus for splitting the lithium nucleus

Fig. 2 The power converter of the Cockcroft and Walton DC generator
(a) Basic principle using switches
(b) Use of diodes and triodes or thyratrons to replace switches
(c) Final circuit using only diodes
Figure 1 shows the original apparatus, which is now kept in the Science Museum, London. The top electrode contains the proton source and was held at 400 kV, the intermediate drift tube at 200 kV and final drift tube and target at earth potential. This structure can be seen inside the evacuated glass tube in Fig. 1 above the curtained booth in which the experimenter sat while watching the evidence of nuclear disintegrations on a scintillation screen. However, the participants in this course will recognise Fig. 2 as being the more important part of the apparatus. In fact, this accelerator was basically a high-voltage "power converter" and it was in this domain that the main technological advance had been made. The principle of operation is shown in Fig. 2(a). The source voltage is first used to charge capacitor C1. The switches are then changed so that C1 can share its charge with C2. On the next cycle, the switches are returned to their initial positions, where the source voltage can top up the charge on C1 and allow C2 to share its charge with C3. On subsequent cycles the charge sharing is carried out to the top of the capacitor system and with further cycles the full output voltage is built up to a multiple of the source voltage. In principle, the number of capacitor stages can be increased to give any multiple of the source voltage. The current that can be drawn from such a circuit depends on the switching rate. This and other practical considerations make it necessary to replace the mechanical switches by a combination of diodes and triodes as in Fig. 2(b) or just diodes as in Fig. 2(c). Cockcroft and Walton had difficulty in obtaining triodes that would stand 200-400 kV, so they chose the all-diode circuit. Even so, the diodes and capacitors were at the limit of their in-house technology. The original design had been for 800 kV and they reached about 700 kV where they were limited by a spark discharge, but the famous atom-splitting experiment was carried out at the fairly modest voltage of 400 kV.

Today, of course, voltages of 400 kV and far higher are used for power transmission, but in 1932 this was a major achievement. The Van de Graaff electrostatic generator [3] (see Fig. 3), which was being developed at the same time, was soon to exceed this level and reach more than 10 MV. In fact, a modified design called the Tandem Van de Graaf operates routinely at the Oak Ridge National Laboratory with 24.5 MV on the central terminal. However, the Cockcroft-Walton Generator, as it became known, was still widely used for many years as the input stage to large accelerators, since it could deliver a much higher current and only in recent years has the Radio-Frequency Quadrupole (RFQ) started to replace it. The RFQ was first suggested in 1970 [4], but it was not demonstrated to work until 1979 [5] at the Los Alamos National Laboratory.

![Fig. 3 Principle of the Van de Graaff electrostatic generator. (By enclosing the whole system in a high pressure tank containing dry nitrogen or Freon at 9-10 atmospheres the sparking threshold can be raised to ~25 MV in a variant called the Tandem Van de Graaff generator.)](image-url)
2. **NEW ACCELERATORS AND THE DEVELOPING ROLE OF THE POWER CONVERTER**

The Cockcroft-Walton DC generator and the Van de Graaff electrostatic generator were only able to accelerate particles to a potential equal to the maximum voltage in the system. Acceleration techniques, which were not limited in this way, had already been proposed by Ising [6] and by Wideröe in an unpublished notebook [7], but the exploitation of their ideas was slower partly due to the technology available.

The differences between the acceleration mechanisms of Cockcroft and Walton, Ising and Wideröe depend upon whether the fields are static (conservative) or time-varying (non-conservative), and in a less fundamental way upon the geometry. The electric field can be expressed in a very general form as the sum of two terms, the first being derived from a scalar potential and the second from a vector potential,

\[ \mathbf{E} = - \nabla \phi - \frac{\partial}{\partial t} \mathbf{A} \quad (1) \]

where the vector potential satisfies the relation

\[ \mathbf{B} = \nabla \times \mathbf{A} \quad (2) \]

This may seem a little theoretical but it soon reduces to more familiar concepts.

The first term in (1) describes the static electric field (or conservative field) of the Cockcroft-Walton and Van de Graaff machines. When a particle travels from one point to another in an electrostatic field, it gains energy according to the potential difference, but if it returns to the original point, for example, by making a full turn in a circular accelerator, it must return to its original potential and will lose exactly the energy it has gained. Thus a gap with a DC voltage has no accelerating effect in a circular machine.

The second term describes the time-varying field. This is the term that makes all the present-day high-energy accelerators function. The substitution of (2) into (1) yields Faraday's Law,

\[ \nabla \times \mathbf{E} = - \frac{\partial}{\partial t} \mathbf{B} \quad (3) \]

which relates the electric field to the rate of change of the magnetic field. There are two basic geometries used to exploit Faraday's Law for acceleration.

**2.1 The "Betatron"**

It was Faraday's Law and the transformer, that led Wideröe [7] to suggest the acceleration mechanism, now known as "betatron acceleration". Wideröe called his idea a "strahlung transformator" or "ray transformer", because the beam effectively formed the secondary winding of a transformer (see Fig. 4). As the flux through the magnet core is increased, it induces an azimuthal e.m.f. which drives the charged beam particles to higher and higher energies. The trick is to arrange for the increase in the magnetic field in the vicinity of the beam to correspond to the increase in particle energy, so that the beam stays on the same orbit. This device, the betatron, is insensitive to relativistic effects and was therefore ideal for accelerating electrons. The betatron has also the great advantages of
being robust and simple. The one active element is the power converter that drives the large inductive load of the main magnet. The focusing and synchronisation of the beam energy with the field level are both determined by the geometry of the main magnet.

2.2 RF linac

In 1924 Ising [6] suggested accelerating particles with a linear series of conducting drift tubes and Wideröe built a "proof-of-principle" linear accelerator in 1928 [8]. Alternate drift tubes are connected to the same terminal of an RF generator. The generator frequency is adjusted so that a particle traversing a gap sees an electric field in the direction of its motion and while the particle is inside the drift tube the field reverses so that it is again the direction of motion at the next gap. As the particle gains energy and speed the structure periods must be made longer to maintain synchronism (see Fig. 5).

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Fig. 4 Strahlung transformer or betatron

Fig. 5 RF linac
At high energies, the drift tubes become very long unless the frequency can be increased. It is therefore natural to enclose the gaps in cavities and to work in the MHz to GHz frequency range. The proposed linear electron colliders for the next generation of accelerators may work at 10’s of GHz. The technology for high-frequency power sources held back the development of linear accelerators until the development of radar in World War II. These sources are basically power converters, but they do not naturally fit into the framework of this course.

Figure 6 indicates the main features of the fields in an RF accelerating cavity. The azimuthal magnetic field is concentrated towards the outer wall and links the beam. Faraday’s Law tells us the periodic rise and fall of this magnetic field induces an electric field on the cavity axis, which can be synchronised with the passage of the beam pulse. By comparing Figs. 4 and 6, it can be seen that the topologies are equivalent. In both cases, the beam and the magnetic field are linked, but for the betatron it is natural to think of the beam linking the field and for the cavity, the field linking the beam.

\[
\oint E_s \cdot d\mathbf{A} = - \int_{\text{cavity}} \frac{\partial B_o}{\partial t} \cdot d\mathbf{A}
\]

Fig. 6 Principal fields in a cavity

The most widely used high-energy accelerator today is the synchrotron. The underlying principle of synchronous acceleration was independently suggested by McMillan [9] and Veksler [10]. In the synchrotron, the beam is maintained on a circular path by a magnetic guide field and the acceleration is provided by one or more RF cavities. The functions of the guide field and the accelerating field are independent and have to be synchronised to keep the beam stable. With the invention of alternating gradient focusing by Christofilos [11] and independently by Courant, Livingston and Snyder [12], the guide field became very sophisticated with very many independently controlled magnets. In a machine like the CERN LEP (see Fig. 7) there are hundreds of power converters running just the guide field, as well as many others powering the RF amplifiers, vacuum pumps, detector magnets and almost all other auxiliary equipment. These power converters are of the DC and slow pulse type, which are the main subject of this course and they fall very broadly into the three categories as indicated in Fig. 8.
Fig. 7 CERN LEP electron-positron collider.
<table>
<thead>
<tr>
<th>High-voltage, low-current</th>
<th>- for klystrons in RF system</th>
</tr>
</thead>
<tbody>
<tr>
<td>Low-voltage, high-current</td>
<td>- for the magnetic guide field (very high to low power)</td>
</tr>
<tr>
<td>Very-low-voltage, high-current</td>
<td>- for super-conducting magnets in physics detectors. - and for magnetic guide field</td>
</tr>
</tbody>
</table>

Fig. 8 Broad classification of power converters and their applications in synchrotrons

The characteristics of these power converters will be influenced by the types of synchrotron, which can be broadly classified as in Fig. 9.

<table>
<thead>
<tr>
<th>Generic type</th>
<th>Description of sub-types</th>
<th>Example</th>
</tr>
</thead>
<tbody>
<tr>
<td>Accelerator</td>
<td>Fast cycling 10's of Hz</td>
<td>RAL-SNS</td>
</tr>
<tr>
<td></td>
<td>Slow cycling Sub-Hz</td>
<td>BNL-AGS</td>
</tr>
<tr>
<td>Storage ring and collider</td>
<td>DC only</td>
<td>CERN-AA</td>
</tr>
<tr>
<td></td>
<td>DC with ramping (over minutes)</td>
<td>CERN-ISRF</td>
</tr>
<tr>
<td></td>
<td>DC with slow cycling</td>
<td>CERN-SppbarS</td>
</tr>
</tbody>
</table>

Fig. 9 Broad classification of synchrotrons

2.3 Principal characteristics of power converters

Figure 10 summarises the characteristics of a power converter that are directly related to the accelerator beam and that are likely to be given a tight specification. Of course, reliability, immunity to mains spikes etc. are no less important but are not dependent on special aspects of the beam optics. Storage rings and colliders will be particularly demanding on stability and ripple, whereas for fast and slow cycling accelerators it will be the "following error".

![Graph showing the characteristics of a power converter](image)

Fig. 10 Principal characteristics
3. **Relation Between Performance, Tolerances and Specifications**

In general, accelerator designers are not power converter experts and vice versa. The interface between the two groups is the list of power converter specifications. The apparent belief on one side that anything is possible and the impression on the other side that the specifications have been chosen at random, sometimes leads to rather cynical feelings such as:

- if in doubt, ask for an extra order of magnitude, there is never any problem, or
- whatever it is, or however it is calculated, 1 in $10^4$ will be good enough in the end.

In addition, the determination of specifications is not well documented in the literature, so it is not easy for those who are interested in teaching themselves. It is usually a matter of visiting the designers in various accelerator projects and discussing with them to find out how they have determined their specifications and obtaining copies of their internal reports.

There should be three main stages in defining specifications:

- Agreeing on **performance** characteristics for the accelerator
- Calculating **tolerances** on the accelerator **parameters**
- Fixing **specifications** for **sub-systems** such as the power converters.

The performance characteristics for the accelerator will be thought of in rather general terms. The maximum beam energy is an obvious example, but there are many less obvious and less well-defined parameters, such as the absolute precision of the beam energy, the positional stability of the beam, the loss rate, the rate of emittance blow-up and so on. In general, these performance parameters will depend upon many sub-systems, for example see Fig. 11.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Sub-system</th>
</tr>
</thead>
<tbody>
<tr>
<td>Loss rate depends on:</td>
<td>- vacuum (scattering on residual gas)</td>
</tr>
<tr>
<td></td>
<td>- main field (stability against collimators)</td>
</tr>
<tr>
<td></td>
<td>- focusing field (loss on resonances)</td>
</tr>
<tr>
<td></td>
<td>etc.</td>
</tr>
</tbody>
</table>

Fig. 11 Example of parameter to sub-system dependence

Similarly, a given sub-system will in general affect many parameters, for example see Fig. 12.

<table>
<thead>
<tr>
<th>Sub-system</th>
<th>Parameter</th>
</tr>
</thead>
<tbody>
<tr>
<td>Main field effects:</td>
<td>- loss rate</td>
</tr>
<tr>
<td></td>
<td>- injection efficiency</td>
</tr>
<tr>
<td></td>
<td>- determination of absolute beam energy</td>
</tr>
<tr>
<td></td>
<td>etc.</td>
</tr>
</tbody>
</table>

Fig. 12 Example of sub-system to parameter influence
The physics behind these relationships is complicated by some rather fundamental differences between electrons, protons and heavy ions. For instance an electron-positron collider such as LEP can tolerate large injection errors for both the electron and positron beams. Providing the beam does not hit something, then after a short while, it will shrink onto the closed orbit due to a mechanism called radiation damping and the final beam size will depend on such things as the coupling due to the tilt errors of quadrupoles and not on the original injection error. In a proton machine, radiation damping is virtually non-existant and the original injection error is "frozen" into the beam size for everafter. Not surprisingly, this difference has a profound effect on the specifications of power converters in proton beam transfer lines and injection systems [13].

Thus the designer has a multi-dimensional problem and he requires considerable understanding of the physics phenomena in order to determine the tolerances on parameters. Basically two approaches are possible. Firstly one can make a list, similar to that started in Fig. 12, of all the parameters influenced by the chosen sub-system. The specification may then be set by the strictest tolerance in the parameter list. For example, the loss rate in a collider is closely linked to the background due to stray particles in the detector. This is likely to impose the tightest tolerance on the current ripple, while the injection efficiency could well determine the resolution of the current setting. Conversely, the ripple would be small compared to the resolution and therefore of no consequence for injection and the resolution at the maximum current would automatically be several times better than at injection and probably quite adequate for making energy scans.

For the second approach it is necessary to work in the reverse sense and to list all the sub-systems contributing to a given parameter, as in Fig. 11. This occurs whenever a particularly tight tolerance has to be satisfied for say the determination of the absolute energy, the luminosity or some equally global parameter.

These two approaches could be classified as:
- the single, strictest criterion, and
- the collective criterion.

4. **SPECIFICATION OF POWER CONVERTERS**

4.1 **Some simple examples**

4.1.1 **Main dipole field**

The short and long term stabilities, for the main dipole field are usually set by the desired positional stability of the beam. The beam optician can calculate how far the beam will move radially at any point in the accelerator for a given field change. Typically this will be 1 to 5 mm per mil of field change. It can be larger, zero, or even negative, but for the present purpose 5 mm per mil of field change will be assumed as a typical maximum.

Figure 13 shows the vacuum chamber cross section of LEP and the corresponding beam size at 60 GeV. It is desirable that the beam stability be a small fraction of the beam size, so it is not surprising to find stabilities of 1 in $10^5$ quoted.
4.1.2 Main quadrupole strings

The main quadrupoles determine the Q of the accelerator, i.e. the number of oscillations that a beam particle will make in one turn. As in most oscillating systems, there are resonance conditions and in the case of alternating gradient synchrotrons there is an infinite number of resonances stretching to infinitely high order. It is not possible to avoid all resonances. First of all, the beam must be positioned between the stronger low-order resonances. Depending on the accelerator optics and whether it is a storage ring or not, this will need a precision in the setting of Q between 0.01 and 0.001. There is not a 1 to 1 correspondence between Q and the main quadrupole currents, but typically this means a precision of 1 in $10^4$ to 1 in $10^5$. In electron rings, however, this may not be sufficient. The radiation damping is characterised by the damping partition numbers and in order that these remain constant to the required degree a further tolerance must be evaluated for the quadrupoles.

Although the stronger resonances can be avoided there are always higher-order resonances inside the beam region. The current ripple causes a Q ripple and these higher-order resonances are moved back and forth inside the beam. This agitation helps to feed particles into the resonances. They are then lost making an unwelcome increase in the background. This is often the tightest tolerance on the ripple in proton colliders and is usually 1 in $10^6$ or better. In a plain storage ring the background is unimportant and a higher loss rate can be accepted making it possible to relax on the ripple tolerance.
4.2 Some more sophisticated examples

4.2.1 Main dipole field ripple

In the mid-sixties the extremely tight tolerance of 1 in $10^7$ was imposed on the current ripple of the ISR main magnet. Two criteria were considered when determining this tolerance.

The first was a beam stacking programme which used very low voltages on the RF cavities and the second was based on the positional stability needed for a slow extraction of the beam onto an external target over several hours. The criterion based on the RF voltage proved to be stricter and can be understood from the explanations given in Section 2 of the two configurations for accelerating a beam, i.e. betatron acceleration and cavity acceleration. In all circular machines, the guide field provides some betatron acceleration as the field increases, although this is normally considered as negligible compared to the contributions from the RF cavities. In the ISR, the guide field was kept constant during stacking, but there was still a minute effect from the current ripple. However, the extremely low cavity voltage required by this particular programme still made it necessary to limit the guide field variations to less than 1 in $10^7$.

As it happened, this stacking programme was never used, and neither was the slow extraction, but fortunately the 1 in $10^7$ tolerance on the ripple was still implemented, since below this level there was an appreciable effect on the background. This was probably due to the fact that the magnets were of the combined function type (i.e. dipole and quadrupole combined) and the current ripple caused a Q ripple which enhanced the losses on resonances. Subsequent experience with the storage of protons and antiprotons in the SPS has confirmed the need for such strict tolerances on the lattice quadrupoles to keep background to a minimum.

4.2.2 Magnet cycling and setting

When the main field setting was discussed earlier, it may have seemed quite straightforward, but here are two more problems to take into account.

Good reproducibility saves time when restarting an accelerator or changing between different working conditions. The hysteresis effect in the iron-cored magnets are the first obstacle in obtaining the level of reproducibility required. Hysteresis loops depend on the steel used, the magnet gap and the maximum field level to which the magnet has been excited. Practical values go up to $\pm 30$ G for the remanent field, which sets the width of the loop. Consider a 'C'-type dipole operating at $\sim 1$ T with an 80 mm gap. The hysteresis loop would be $\sim 20$ G or $\pm 2 \times 10^{-3}$ of the maximum field. This is a large error and the usual solution is to cycle the magnet in a pre-determined way before setting. With care, the field uncertainty can be brought down to $\pm 0.5$ G or $\pm 5 \times 10^{-5}$ of maximum.

Now, what does "with care" mean? Figure 14 shows a typical magnet-setting cycle, (a) is the hysteresis loop for the cycle and (b) is the current-time profile. In part (b), there are reduced ramp rates just prior to the minimum of the cycle and the set value [14]. These are intended to prevent overshoot. Overshoot at the top of the cycle is less critical, especially if the magnet steel is fully saturated, so the reduced ramp rate has been missed out at that point. Figure 15 shows the effect on the hysteresis of overshoot when approaching the final setting. Note that the sub-loop
set-up by the overshoot never returns to the ideal value. The only sure way to correct the error is to re-cycle. Thus overshoot is important and at the level of $10^{-5}$ of maximum.

![Hysteresis loop](image1) ![Current-time profile](image2)

(a) Hysteresis loop (b) Current-time profile

Fig. 14 Typical magnet setting cycle.

![Effect of overshoot](image3)

Fig. 15 Effect of overshoot on the field setting

4.2.3 Ripple and frequency effects

Current ripple is characterised by its amplitude and frequency. For all practical purposes the beam responds adiabatically to the ripple, that is the ripple is slow enough (i.e. compared to the revolution frequency) that the beam can adjust itself continuously, so that its emittance remains constant (i.e. its area in phase space). The magnets and the vacuum chamber are more sensitive to the frequency and progressively screen out the higher frequencies. In the ISR the magnets were equipped with poleface windings and in order to protect these windings, when baking the vacuum chamber, they were covered by a thin, copper, heat shield. The attenuation (see Fig. 16) due to this heat shield made it possible to considerably relax the ripple tolerance at higher frequencies [15]. A more subtle problem is due to the fact that large synchrotron magnet systems have appreciable capacitance to earth and to all except very low frequencies look like transmission lines. Reference 16 investigates the standing waves set up by an abrupt voltage change and the effect of ripple making the current a function of azimuthal position. In both cases, there is a closed orbit distortion and a tolerance can be determined based on aperture.
5. CONCLUSION

It is hoped that these few examples will illustrate the very broad approach that is needed when determining tolerances and the specifications of power converters. The aim was not to list all the possible criteria that might be useful in determining a specification. Such a list would never be complete and there is no fixed recipe that will always determine the optimum values. The aim was rather to illustrate that the apparent randomness in specifications between different machines has a sound scientific basis. When determining tolerances, it is essential to have a detailed understanding of the physics behind the operation of the accelerator and of the special requirements in a particular case.

For the foreseeable future, accelerator energies will rise and beam sizes will decrease, which will no doubt mean even tighter tolerances for the power converter designer.
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CLASSIFICATION OF STATIC CONVERTERS

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1. INTRODUCTION

Particle accelerator installations require large direct current power supplies for the magnets, klystrons, vacuum pumps and other equipment [1-3]. For this purpose, static converters are used to convert alternating current (a.c.) to direct current (d.c.).

These power supplies are required to provide precision voltage and current regulation over a wide range with low ripple and high stability. Large smoothing filters and high precision regulating circuits must be used to meet these requirements.

I shall first briefly introduce four of the most commonly used static converters in this field. For the layman, I shall review the fundamentals of conversion techniques and semiconductors. After some general observations about precision and dynamic performance, I shall describe the configurations and operation of the four static converters, i.e. the current converter, three-phase current regulator and diode rectifier, chopped d.c. regulator and a d.c. regulator with an intermediate oscillating circuit.

2. STATIC CONVERTERS

2.1 General

Essentially four different types of static converters are used for d.c. power supplies in accelerator installations. These are shown in block diagram form in Fig. 1.

![Fig. 1 Static converters for power supplies in accelerator installations; (a) current converter (b) three phase current regulator and diode rectifier (c) chopped d.c. regulator (d) d.c. regulator with an intermediate oscillating circuit.](image-url)
Figure 2 shows voltage $U_d$, current $I_d$ and power $P_d$ for these four types of static converters. These are only indicative limits. Current developments in semiconductor power components will enable these boundaries to be extended in the future.

2.2 Current converter

The current converter in Fig. 1(a) comprises a transformer 1, a thyristor bridge 2 and a smoothing filter 3. Transformer 1 provides both galvanic isolation and adjustment of the level of the a.c. voltages. The thyristor bridge converts a.c. to d.c. and regulates the d.c. voltage (controlled rectifier). The filter 3 is needed to smooth the voltage harmonics produced by the current converter.

The d.c. voltage is several hundred volts and the d.c. current several thousand amps.

2.3 Three-phase current regulator and diode rectifier

The three-phase current regulator and diode rectifier (Fig. 1(b)) converts a.c to d.c. and regulates the voltage in two stages. Transformer 1 provides galvanic isolation and matches the voltage level to the three-phase current regulator 2. It comprises thyristors in antiparallel, i.e. head-to-tail, and allows the three-phase voltage to be regulated. Transformer 3 matches the voltage level to the d.c. output voltage. Diode rectifier 4 converts a.c. to d.c. Smoothing filter 5 reduces the ripple caused by rectification.

This type of static converter lends itself well either to very high d.c. voltages with low currents or for very high currents with low d.c. voltages. These configurations allow the thyristors and diodes to be used in the best way.

2.4 Chopped d.c. regulator

The chopped d.c. regulator also converts a.c. to d.c. and regulates voltage in two stages. Transformer 1 provides a galvanic separation and matches the voltage level to the output d.c. voltage. It powers diode rectifier 2.
Smoothing filter 3 smooths the rectified voltage. The d.c. regulator 4, comprising switching transistors, adjusts the voltage on the pulsation or chopper principle. A smoothing filter 5 is also required to smooth the chopped d.c. voltage.

This type of static converter can at present be produced for some tens of kW of power.

2.5 The d.c. regulator with an intermediate oscillating circuit

The d.c. regulator with an intermediate oscillating current (resonance converter) shown in Fig. 1(d) converts a.c. to d.c. and regulates the voltage in three stages. As for Fig. 1(c), there is a transformer 1, a diode rectifier 2 and a smoothing filter 3. The fixed d.c. voltage is applied to inverter 4 which is fitted with switching transistors. It operates at high frequency, powers the series oscillating circuit 5 and regulates the voltage. Transformer 6 adjusts the voltage level while rectifier 7 supplies the d.c. output voltage. A capacitor 8 is all that is needed to smooth this voltage owing to the operation as an r.f. current source. This type of static converter is produced at present for some tens of kW of power.

3. CONVERSION METHODS

3.1 General

The operation of static converters is based on the fundamental principle of the conversion technique: periodic switching between different voltage levels. According to the voltage levels and the sequence of the switching selected, different types of conversion may be obtained. Figure 3 shows in block diagram form the conversions used in the static converters referred to in section 2.

![Diagrams showing conversion methods](image)

Fig. 3 Conversion methods: (a) a.c.-d.c. (b) a.c.-a.c. at fixed frequency (c) d.c.-d.c. (d) d.c.-a.c.

The switching principle makes it possible to use semiconductor power elements operating only in two modes: conducting or blocked. In this way, losses are greatly reduced and the static converters give a fairly good performance. On the other hand, varying ripple or voltage harmonics occur which have to be reduced by smoothing filters in the power circuits.
3.2 a.c.-d.c. conversion

For a.c.-d.c. conversion, (Fig. 3(a)), periodic switching is made between sinusoidal sectors in a multiphase a.c. system. The average value of the d.c. voltage is proportional to \( \cos \alpha \) and may therefore be regulated over a wide positive and negative range. The number of phases \( p \), also termed the pulse number, determines the size of the ripple and its frequency. The latter is \( f_0 = pf \), where \( f \) is the frequency of the a.c. power supply. The thyristor current converter and the diode rectifier work on this principle (in the case of the latter, \( \alpha \) is always 0).

3.3 a.c.-a.c. conversion at a fixed frequency

For an a.c.-a.c. conversion at a fixed frequency (Fig. 3 (b)), switching is done symmetrically in the two half-waves between a sinusoidal voltage sector and zero. Using the angles \( \alpha \) and \( \beta \), the fundamental wave \( \hat{U} \) of the a.c. voltage can be regulated between \( \hat{U} \) and 0. In the single-phase case, the pulse number \( p = 2 \), and for three phase \( p = 6 \). The size of the harmonics also depends on \( \alpha \) and \( \beta \). Their frequency is a whole multiple of the power supply frequency \( f \).

This conversion principle is found in the a.c. regulator (single-or three-phase).

3.4 d.c.-d.c. conversion

d.c.-d.c conversion (Fig. 3 (c)) involves switching between two constant d.c. voltages (e.g. +U and -U or +U and 0). The duration of switch-on, \( t_c \), and switch-off, \( t_d \), determines the average value \( U_d \) of the d.c. voltage. It may be positive or negative (bidirectional) for switching between +U and -U. It is positive only (unidirectional) for switching between +U and 0. The size of the ripple depends on \( t_c \) and \( t_d \). The frequency is \( f_0 = 1/T_p \), where \( T_p \) is the pulse period.

This conversion principle is used for the chopped d.c. regulator.

3.5 d.c.-a.c. conversion

There are several methods for d.c.-a.c. conversions. The simplest involves symmetrical switching between constant voltages +U and -U and zero (Fig. 3 (d)). If \( t_c \) is varied between 0 and \( T/2 \), the amplitude \( \hat{U} \) of the fundamental wave can be varied between 0 and \( 4U/\pi \). The harmonics depend on the ratio between \( t_c \) and \( T/2 \) and their frequency is a multiple of \( f = 1/T \).

This conversion principle is the basis of the inverter.

4 SEMICONDUCTOR POWER ELEMENTS

4.1 General

In static converter configurations, semiconductor power elements are used as static switches. The components used in this context are summarized in Fig. 4 with their symbols and characteristics. Depending on the specific voltages and the currents, a number of semiconductor components may have to be put in series or in parallel.
4.2 Diode

The diode is a semiconductor component that cannot be controlled. It conducts the current in the positive direction (anode cathode) and blocks in the negative direction (cathode anode).

4.3 Thyristor

The thyristor is a semiconductor component that can be controlled at turn-on. In principle it blocks the current in the positive and negative directions. It conducts the current in the positive direction only if a firing pulse is given between the trigger and the cathode provided that the voltage between the anode and cathode $U_{AC}$ is $> 0$. The thyristor blocks when current $I$ goes to zero, provided that $U_{AC} < 0$ for a period of time longer than $t_q = 100$ to $200$ $\mu$s (for slow thyristors) or $t_q = 5$ to $50$ $\mu$s (for high-speed thyristors) after the current reached zero.

4.4 Bipolar transistor

The bipolar transistor is a semiconductor component that can be controlled at turn-on and turn-off by means of the base current $I_B$. In the conducting mode ($I_B > 0$), the transistor is completely saturated and it conducts a positive current $I$. When blocked ($I_B = 0$), the voltage $U_{CE}$ between the collector and the transmitter must be positive. It should be noted that the transistor cannot withstand negative $U_{CE}$ voltages.

At turn-on and even more so at turn-off, losses occur which may limit the switching frequency.
4.5 MOSFET Transistor

The MOSFET (Metal Oxide Semiconductor Field Effect Transistor) is also a semiconductor component that can be controlled at turn-on and turn-off by means of the voltage $U_{GS}$ between the gate and the source. In conducting mode ($U_{GS} > 0$) the MOSFET is completely saturated and it conducts a positive current $I$. In blocked mode ($U_{GS} = 0$), the voltage $U_{DS}$ between the drain and the source must be positive. It should be noted that the MOSFET conducts in the negative direction ($I < 0$) due to its physical structure.

This type of transistor generally tolerates higher switching frequencies than the bipolar transistor.

4.6 "Dual" thyristor

The "dual" thyristor consists of a transistor (fitted with an appropriate control linked to a control signal and to the voltage at the transistor terminals), a capacitor and a diode in antiparallel [4]. It shows duality with a thyristor (i.e. currents are replaced by voltages and vice versa, and the conducting and blocked modes are inverted). Consequently, the "dual" thyristor can be controlled when blocked (turn-off). In principle it conducts the current in both the positive and negative directions. The "dual" thyristor blocks the current in the positive direction only if the appropriate control signal is given, where $U > 0$. The capacitor limits the $dU/dt$ in the blocked mode which considerably reduces switching losses. The "dual" thyristor allows a high frequency inverter to be installed in a d.c. regulator with an intermediate oscillating circuit.

5. PRECISION AND DYNAMIC PERFORMANCE

5.1 General

Very high current precision is required to ensure worthwhile results in physics experiments [2], where tolerance levels are no more than $10^{-4}$ of the maximum d.c. value. Current fluctuations must remain within this tolerance range irrespective of the disturbances that may occur. There are generally three types of disturbances defined according to their frequency bands:

- high speed disturbances (300 Hz to 100 kHz) due to voltage ripple of the static converters;
- fast disturbances (0.1 to 50 Hz) due to small fluctuations in voltage in the three-phase mains;
- slow disturbances ($10^{-5}$ to $10^{-2}$ Hz) due to heating up of components and magnets; in this context, reproducibility from one experiment to the next is important.

The negating effects of ripple can only be reduced by the means of active or passive smoothing filters. In order to overcome the effect of fast and slow interference, regulating circuits are needed. In this context the dynamic behaviour of static converters is significant.

5.2 Static converter ripple

In order to estimate the extent of load current ripple and hence to establish the dimensions of the smoothing filter, the maximum ripple of the d.c. voltage supplied by the static converter must be known. The voltage
depends on the operating point. Figure 5 shows the relationships for the static converters in question and the peak-to-peak value $U_d'$ of the ripple is also given referred to the maximum value $U_{d\text{max}}$ of the d.c. voltage as a function of the frequency $f_0$ of the ripple.

![Diagram showing the relationships for static converters](image)

**Fig. 5**  d.c. voltage ripple; (a) current converter (b) three-phase current regulator and diode rectifier (c) chopped d.c. regulator (1: unidirectional; 2: bidirectional) (d) d.c. regulator with an intermediate oscillating circuit

Current converter ripple (Fig. 5a) depends on the pulse number $p$. The maximum value appears where $\alpha = 90^\circ$ [5]. It is relatively low but its frequency $f_0$ is also very low (between 300 and 1200 Hz as a function of $p$). It should be noted that similar ratios are obtained for the three-phase current regulator and diode rectifier (Fig. 5b).

The ripple of a current regulator is most pronounced where $t_c = t_d = T_p/2$ (Fig. 5c). It is relatively high and depends on whether the configuration is unidirectional ($U_d$ positive) or bidirectional ($U_d$ positive and negative). On the other hand, the frequency $f_0 = 1/T_p$ is quite high. Nowadays, values up to 50 kHz may be reached.

The d.c. regulator with an intermediate oscillating circuit has a special feature. As a current source, it has a current ripple rather than a voltage ripple (Fig. 5d). The ripple is made up of half-waves of sinusoidal shape. The frequency $f_0 = 2f = 2/T$ is very high. The frequency $f$ in the intermediate oscillating circuit may reach values between 20 kHz and several hundred kHz.

5.3 **Passive smoothing filters**

Passive smoothing filters are LC circuits as shown in Fig. 6(a). There are two smoothing self-inductors (induction coils) $L_f$ placed in the two lines and a capacitor $C_f$ in parallel with the ohmic and inductive charge $R_{ch}$. 
L\(_{ch}\) (magnet). It should be noted that the dimensions of the smoothing filter must be matched to the power absorbed by the magnet. Figure 6 (b) shows the modulus of the harmonic response of \(G(s) = \frac{i_{ch}}{u_d}\) which clearly shows the cut-off frequency \(1/T_{ch}\) due to the time constant \(T_{ch} = L_{ch}/R_{ch}\) of the load (magnet), and the cut-off frequency (resonant frequency) \(\omega_f = 1/2\sqrt{L_{ch}C_f}\) of the smoothing filter. Adequate damping with resistors \(R_f\) and \(R_c\) must be provided to avoid the negative effects of the resonance.

![Diagram](image)

Fig. 6  Passive smoothing filter: (a) circuit layout (b) harmonic response

The resonant frequency \(f_r\) must be less than the ripple frequency, \(f_0\), to obtain the necessary attenuation of the current ripple. For the d.c. regulator with an intermediate oscillating current, the inductors \(L_f\) must be removed and only the capacitor \(C_f\) kept. Given that the size and cost of a smoothing filter increase as the resonant frequency \(f_r\) decreases, it can be seen that the chopped d.c. regulators and those with an intermediate oscillating circuit are more economical than the current converter. Unfortunately, these regulators can be produced only for a relatively restricted power range (see Fig. 2).

5.4 Active smoothing filters

Sometimes it is worthwhile substituting active smoothing filters for passive filters as shown diagrammatically in Fig. 7.

![Diagram](image)

Fig. 7  Active smoothing filter

Using two transformers \(T\), an amplifier \(A\) induces a.c. voltages to offset the d.c. voltage ripple supplied by the static converter. This amplifier, composed of continuously operating transistors, must be dimensioned for relatively low power, i.e. that given by the voltage and current ripple (the current ripple must be very low, about \(10^{-5}\) of \(I_{d_{max}}\)).

5.5 Control circuits

To ensure the required precision in the event of fast and slow disturbances, the static converters must be provided with control circuits. Figure 8 shows the block circuit diagram [6,7]. As can be seen, cascade control is used [8]. Static converter 1 powers magnet 3 or other consumers (e.g. a klystron) via filter 2. The static converter's d.c. voltage can be adjusted by the control signal \(u_{cm}\) via control unit 4 (a firing control for the
thyristor, or control for bases of the bipolar transistors or MOSFETs). An initial control loop is formed by voltage regulator 5. The d.c. voltage \( u_d \) is measured downstream of the smoothing filter by means of d.c. voltage transformer 6. The reference value \( u_c \) comes from super-imposed current regulator 7. The d.c. \( i_{ch} \) circulating through the load (magnet) is measured by d.c. transformer 8. The reference value \( i_c \) for the current is derived from the reference generator 9.

![Control circuit block diagram](image)

Fig. 8 Control circuit block diagram

The voltage control circuit is designed to reduce the effects of fast disturbances. It must therefore have good dynamic performance. Voltage regulator 5 may be high-speed analog or digital. Since the smoothing filter 2 enters into the control circuit, it may be advisable to use a state feedback. Better dynamic performance is then obtained than with a standard PI type regulator [6]. On the other hand, the current control circuit has to combat slow disturbances and above all guarantee highly accurate reproducibility. For this, a digital control for the current regulator 7 and the reference generator 9 is required, mainly to avoid the drift of operational amplifiers. D.c. measurements (d.c. transformer 8) must also be very accurate and stable.

5.6 Dynamic behaviour of static converters

The dynamic behaviour of the static converter and its command circuit enter into the voltage control loop. Depending on the conversion technique used, based on periodic commutations, the d.c. voltage can be modified only at regular intervals, which introduces an element of delay. This delay can be expressed approximately by a small time constant \( T_{cm} \). Its value may be determined either by consideration as an average dead time [5], or by treating it as a sampled system [9].

For a current converter or a three-phase current regulator, this small time constant is \( T_{cm} = T/2p \), where \( T \) is the period of the three-phase mains powering the static converter and \( p \) is the pulse number. For a chopped d.c. regulator, \( T_{cm} = T_p/2 \) where \( T_p \) is the period length. The dynamic behaviour of a d.c. regulator with an intermediate oscillating circuit is fairly complex. As a first approximation, a small time constant \( T_{cm} = 1/2f \) may be used in the calculation where \( f \) is the frequency of the intermediate oscillating circuit.

Figure 9 shows the values of the small time constant for various static converters. It should be noted that the regulation speed is limited by the time constant \( T_{cm} \) of the control unit [8]. To obtain high-speed regulation, a low time constant \( T_{cm} \) is needed. In this respect the current converter is equally disadvantageous. However, it has to be used when the required continuous power is relatively high. The three-phase current regulator and diode rectifier has the same disadvantage.
6. **CURRENT CONVERTER**

6.1 **General**

As we have said, the current converter is widely used and up to high power levels [10]. Its configuration and operation will be described in more detail below. Its effects on the mains, and the measures needed to increase the pulse number, will also be mentioned.

6.2 **Three-phase bridge configuration**

![Three-phase bridge current converter](image)

A transformer with its coils mounted in star/star configuration is connected to a current converter comprising six thyristors. This is the three-phase bridge configuration shown in Fig. 10. This feeds the load \( R_{ch}, L_{ch} \) (magnet) with variable direct voltage and current via the smoothing filter \( L_f, C_f \). Sometimes, a diode \( D_R \), also known as a free-wheel diode, is connected to the thyristor bridge in parallel. This avoids instantaneous negative values for the d.c. voltage \( u_{id} \), reduces voltage ripple and enables the energy of the magnet to be discharged in the event of a mains failure.

6.2.1 **Operation**

Normally, the d.c. current passes through two thyristors (one in each half of the bridge). The instantaneous value of the d.c. voltage \( u_{id} \) is therefore identical to that of the voltage between the two phases conducting the current. No current is conducted on the third phase. When the current is switched from one thyristor to another (separately in each half of the bridge), there is for a brief moment a short-circuit between two phases. These switching operations are cyclic. The d.c. voltage \( u_{id} \) is therefore the sinusoidal part of a six-phase system as shown in Fig. 11. This waveform corresponds to a pulse number \( p = 6 \). A highly pronounced voltage ripple can be observed.
Fig. 11  Waveform of the direct voltage $u_d$ and current $i_d$ of a current converter with a three-phase bridge configuration

The firing delay angle $\alpha$ defines the firing time of one of the thyristors. The average value $U_{d\alpha}$ of the d.c. voltage may be regulated by means of this angle. Angle $\mu$ is the overlap angle during which the current is switched from one thyristor to another. This switch causes a drop in the d.c. voltage, known as the inductive voltage drop. At 2-6% of the maximum value of the d.c. voltage, it is relatively low and proportional to $i_d$ and to the short-circuit inductance, $L_C$, of the transformer and the mains. The current $i_d$ can only be positive, given the direction of the thyristors' conductivity. This current has a slight ripple. The ripple depends essentially on the inductance $2L_f$ of the smoothing filter.

6.2.2  Ideal and on-load characteristics

The average value of the d.c. voltage is given by

$$U_{d\alpha} = U_{di0} \cos \alpha - R_1 i_d$$  (1)

where $U_{di0} = 2.339 U_y$ and $R_1 = \frac{3}{\pi} \omega L_C$ with $U_y$ the r.m.s. value of the basic voltage on the secondary of the transformer, $\omega = 2\pi f$ the mains angular frequency and $L_C$ the short-circuit inductance [10].

The ideal ($i_d = 0$) and on-load characteristics of the current converter can therefore be plotted (see Fig. 12). For $\alpha > 90^\circ$ the direct voltage $U_{d\alpha}$ is negative. This range may be used to de-excite the magnet rapidly. To avoid commutation malfunction, $\alpha$ should be limited to about $150^\circ$.

Fig. 12  Characteristics of a current converter (a) ideal (1: without 2: with a free-wheel diode $D_0$), (b) on-load (without a free-wheel diode).
Relation (1) is valid only when there is no diode \( D_4 \) in antiparallel (free-wheel diode). This diode prevents instantaneous negative values of the d.c. voltage \( u_d \). In this case, this phenomenon appears where \( \alpha > 60^\circ \). The ideal characteristic is then modified as shown in the curve 2 of Fig. 12(a).

6.2.3 Effects on the mains

The current converter has effects on the mains as shown in Fig. 13. The mains current (phase current) \( i_1 \) is not sinusoidal but has an almost rectangular waveform. Moreover, the fundamental ripple of this current is out of phase with the basic mains voltage \( u_1 \) by the phase angle \( \phi = \alpha + 120^\circ \). This requires reactive power, despite the d.c. load. Pulses appear on the voltage \( u_1 \) owing to the commutation of the d.c. current from one phase to another.

![Waveform of the mains voltage \( u_1 \) and current \( i_1 \) on the secondary of the transformer](image)

Fig. 13 Waveform of the mains voltage \( u_1 \) and current \( i_1 \) on the secondary of the transformer

Current and voltage harmonics on the a.c. side may have harmful effects on the mains especially in the case of high power current converters. Voltage distortions and disturbances by inductive or capacitive coupling to other users are the consequences. If need be, absorption filters (LC filters) must be fitted between the mains and the transformer to reduce the harmful harmonic effects [10]. Another possibility, i.e. the increase in the pulse number \( p \), will be presented in the next section.

It should be noted that disturbance phenomena also appear on the other types of static converters. Effectively they have a diode rectifier at the input. This works like that of a current converter, where \( \alpha = 0^\circ \).

6.3 Twelve-pulse current converters

The current converter with a three-phase bridge configuration which we have discussed so far has a pulse number \( p = 6 \). As shown in section 5, that has disadvantages in respect of voltage harmonics and dynamic behaviour. In addition, the effects on the mains are very pronounced. The pulse number can be doubled by two three-phase bridges in a series or parallel configuration. This reduces the harmful effects.
Fig. 14 Twelve-pulse current converter, two three-phase bridges in series

Figure 14 shows two three-phase bridges in series. The transformer has two three-phase windings on the secondary which are star and delta configurations respectively. This gives two three-phase systems staggered by 30° as shown in the vector diagram in Fig. 15 (a). Obviously this can be achieved in a number of ways. Often two identical transformers are used, the primary in an asymmetrical zig-zag configuration, giving 15° phase shift [11]. The second transformer is connected to the mains with an opposite phase sequence. At the star-mounted windings on the secondary, two three-phase systems are phase shifted by 30°, as shown in Fig. 15 (b).

(a)  
(b)

Fig. 15 Vector diagram of the transformer voltage (a) star/star and delta connection (b) zig-zag/star connection

When the two three-phase bridges are connected in series, the resulting direct voltage $u_d$ has a pulse number $p = 12$, as shown in Fig. 16 (a). The a.c. $i_P$ on the primary with a stepped waveform closely resembles a sinusoidal waveform (Fig. 16 (b)).

In order to achieve the desired reduction of d.c. voltage ripple and a.c. current harmonics, it is essential for the transformer and the thyristor controls to be precisely symmetrical.

When doubling the pulse number, it is generally not necessary to have any more thyristors than for a three-phase bridge since for high power levels thyristors must be connected in series or in parallel. On the other hand, the transformer design is a little more complicated because of the two windings on the secondary. Its rated power is slightly higher when the zig-zag/star configuration is used.
Fig. 16 Two three-phase bridges in series: the resulting waveforms of the d.c. voltage (a) and the a.c. current (b).

It may be advisable to provide two parallel three-phase bridges for power supplies with high current. This configuration is shown in Fig. 17. This parallel connection must be made downstream of the smoothing filter. The smoothing self-inductors function at the same time as absorption coils [10] and absorb the difference between the instantaneous values of direct voltages $u_{dII}$ and $u_{dIII}$. The effect on the d.c. pulse number and the a.c. waveform is the same as for the series configuration.

Fig. 17 Twelve-pulse current converter, two three-phase bridges mounted in parallel

6.4 24-pulse current converters

A current converter with a pulse number $p = 24$ is obtained by linking four three-phase bridges in series or in parallel. Each must be powered by a three-phase system mutually phase shifted by 15°. Therefore, transformers with zig-zag windings are usually used [11].

Perfect symmetry is even more essential. Consequently, this principle is not suitable for subsequently increasing the pulse number.
7. **THREE-PHASE CURRENT REGULATOR AND DIODE RECTIFIER**

7.1 **General**

Designing a current converter may become problematic where high d.c. voltages (approx. 100 kV) are required, e.g. for klystron power supplies. It may be advisable in such cases to use a more complex configuration using a three-phase current regulator and diode rectifier. This allows more effective use of the thyristor current and voltage ratings. The configuration and operation will be described below as well as methods to increase the pulse number.

7.2 **Configuration**

The configuration of a three-phase current regulator and diode rectifier is shown in Fig. 18. The first transformer T1, e.g. in a star/star configuration, feeds a three-phase current regulator [10]. This comprises two thyristors in antiparallel in each phase. The three-phase current regulator charge is constituted by a second transformer T2 in a delta/star configuration, upstream of a diode rectifier bridge II. The bridge powers the load (e.g. a klystron) via the smoothing filter.

![Fig. 18 Configuration of a three-phase current regulator and diode rectifier](image)

7.3 **Operation**

When describing the operation of this configuration, ideal behaviour is assumed and the overlap in the three-phase current regulator and diode rectifier is ignored. Furthermore, the d.c. $i_d$ is assumed to be completely smooth. The operating principle is shown in Fig. 19. The current $i_R$ in the three-phase current regulator, the voltage $u_{s1}$ and the current $i_{s1}$ on the secondary of transformer T2, as well as the direct voltage $u_d$ are shown.

The firing instant of the thyristors of the three-phase current regulator may be regulated by the firing delay angle $\alpha$. The voltage on the secondary of the transformer T2 depends on the conducting state of the thyristors. If a thyristor conducts in each phase, line voltages are supplied by transformer T1 to the secondary of T2 (taking account of its transformation ratio). Only one diode is conducting in each half of the rectifier bridge, as in the case of the normal operation of a rectifier bridge. On the other hand, if only two phases of the three-phase current regulator conduct, the interconnection voltage is found on one phase of the secondary of T2 and on the other two, half this voltage but with the opposite sign. Two diodes are conducting in one half of the rectifier bridge. If no phase of the three-phase current regulator is conducting, the voltage at the secondary is zero. In that case the rectifier bridge acts as a short circuit on the d.c. side. Consequently, the waveform of voltage $u_{s1}$ and of current $i_{s1}$ can be plotted. The waveform of voltages $u_{s2}$ and $u_{s3}$ and currents $i_{s2}$ and $i_{s3}$ of the other phases is identical to $u_{s1}$ and $i_{s1}$, but displaced by 120° or 240°.
Fig. 19 Waveforms of the current \(i_R\) in the three-phase current regulator, the voltage \(u_{s1}\) and the current \(i_{s1}\) on the secondary of the transformer \(T_2\), and the direct voltage \(u_d\).

The waveform of the d.c. voltage \(u_d\) is obtained by rectifying voltages \(u_{d1}\), \(u_{d2}\) and \(u_{d3}\). As can be seen, d.c. voltage \(u_d\) has a pulse number \(p = 6\). It should be noted that the waveform of \(u_d\) is not the same as that of a current converter. In particular, it can never be negative.

### 7.4 Ideal characteristic

For the ideal characteristic (see Fig. 20) giving the relationship between the average value \(U_{dio}\) and the firing delay angle \(\alpha\) three ranges have to be noted:

\[
\begin{align*}
0^\circ & \leq \alpha \leq 60^\circ & U_{dio} &= \frac{U_{dio}}{2} (1 + \cos \alpha) \\
60^\circ & \leq \alpha \leq 90^\circ & U_{dio} &= \frac{U_{dio}}{2} \sqrt{3} \sin(\alpha + 60^\circ) \\
90^\circ & \leq \alpha \leq 150^\circ & U_{dio} &= \frac{U_{dio}}{2} \sqrt{3} [1 - \cos(\alpha - 150^\circ)] \\
\end{align*}
\]

where \(U_{dio}\) is the maximum d.c. voltage for \(\alpha = 0^\circ\).

Fig. 20 Ideal characteristic of a three-phase current regulator and diode rectifier
To obtain the characteristic on load (as a function of \( i_d \)), it is advisable to make a numerical simulation to take account of the various inductances and resistances that cause voltage drops [12].

7.5 Twelve-pulse configuration

As we have seen, the three-phase current regulator and diode rectifier has a pulse number \( p = 6 \), like a current converter with a three-phase bridge. Here too, the pulse number can also be doubled to obtain \( p = 12 \). Two three-phase current regulators are needed, powered by two three-phase systems phase shifted by 30° which can be obtained at the first transformer. The two diode rectifiers, separately powered by a transformer, can then be connected in series or in parallel. The detailed configuration will not be shown here (see [12]).

8. CHOPPED d.c. REGULATOR

8.1 General

As shown in section 5, the chopped d.c. regulator has undoubted advantages with respect to the size of the smoothing filter and dynamic behaviour. However, it should be noted that this type of static converter can be built only for relatively low power (see Fig. 2). Switch-mode power transistors are used - i.e. either bipolar or MOSFET transistors [13-16]. In this section the configuration and the operation of the chopped d.c. regulator will be described and a distinction made between unidirectional and bidirectional operation [5].

8.2 Configuration

The configuration of a d.c. regulator is shown in Fig. 21. A transformer \( T_1 \) feeds a diode rectifier I which supplies the constant d.c. voltage \( U_e \) via a first smoothing filter. This then is the input voltage for the d.c. regulator II proper. The latter comprises a transistor and a diode in the case of a unidirectional regulator (Fig. 21 (a)). In the case of a bidirectional regulator (Fig. 21(b)), four transistors with antiparallel diodes forming the two

![Diagram](image)

Fig. 21 Configuration of a chopped d.c. regulator: (a) unidirectional  (b) bidirectional
branches of a bridge are needed. The system is often called an H bridge. Since the chopped d.c. regulator switches voltage $u_d$ between two levels (i.e. $U_e$ and 0, or $U_e$ and $-U_e$), another smoothing filter is needed to smooth the load voltage $u_{ch}$ and load current $i_{ch}$.

8.3 Operation

With the unidirectional d.c. regulator $u_d = U_e$ when the transistor is conducting. The diode is blocked. If, on the other hand, the transistor is blocked, the diode conducts the d.c. $i_d$ and $u_d = 0$, as shown in Fig. 22 (a). In this case, only one quadrant in the plane $(u_d, i_d)$ can be used - where $u_d$ and $i_d$ are positive. The average d.c. voltage value $U_d$ depends on the triggering time $t_e$ and switch-off time $t_d$. The pulse period is $T_P = t_e + t_d$.

Fig. 22 Waveform of d.c. voltage $u_d$ and current $i_d$ of a chopped d.c. regulator: (a) unidirectional (b) bidirectional.

With the bidirectional d.c. regulator, the four quadrants can be used, i.e. $u_d$ and $i_d$ can be positive or negative. Depending on whether the thyristors or diodes are conducting, $u_d = U_e$, 0 or $-U_e$.

Figure 22 (b) shows the waveform of voltage $u_d$ for a bidirectional regulator. For this purpose switching between $+U_e$ and $-U_e$ has been assumed. There is a very pronounced voltage ripple. The average d.c. voltage value $U_d$ also depends on the ratio between the triggering time $t_e$ and the pulse length $T_P$. With an appropriate control, the d.c. voltage can also be chopped between $+U_e$ and 0 or $-U_e$ and 0, thereby reducing voltage ripple. The transition of the direct voltage $U_d$ through zero creates some problems.

The d.c. current $i_d$ has a slight triangularly-shaped ripple. It depends essentially on the inductance $2L_f$ of the smoothing filter. The chopping frequency $f_p = 1/T_p$ is limited by switching losses and also by transistor turn-on times $t_{on}$ and turn-off $t_{off}$. Indeed, the whole current $i_d$ has to be switched at turn-on and turn-off.

8.4 Ideal and on-load characteristics

In ideal operating conditions, the average d.c. voltage value $U_d$ is given, for a unidirectional regulator, by the formula
\[ U_{o_1} = \frac{t_s}{t_s + t_d} U_s = \frac{t_s}{T_p} U_s \]

and for a bidirectional regulator by

\[ U_{o_2} = \frac{t_s - t_d}{t_s + t_d} U_s = \frac{2t_s}{T_p} U_s \]

The ideal characteristics can be plotted from these relationships (Fig. 23 (a)).

![Graph showing characteristics of a chopped d.c. regulator](image)

(a)

(b)

Fig. 23  Characteristics of a chopped d.c. regulator: (a) ideal (1: unidirectional 2: bidirectional), (b) on-load (bidirectional).

For relatively high switching frequencies, \( f_p = 1/T_p \), transistor switching times \( t_{\text{on}} \) and \( t_{\text{off}} \) must be taken into account. This creates a lag in the effective switching times as opposed to ideal times. Since \( t_{\text{off}} \) depends on the \( I_d \) to be switched, the lag causes a voltage drop as a function of the current. The on-load characteristic is shown in Fig. 23(b). The characteristic is discontinuous when the current passes through zero.

9. THE d.c. REGULATOR WITH INTERMEDIATE OSCILLATING CIRCUIT

9.1 General

As shown in section 5, high switching frequency is an advantage when considering the size of the smoothing filter and the regulation speed. As has been said, in the case of the chopped d.c. regulator, switching frequency is mainly limited by switching losses. When it is possible to switch with zero current the problem will no longer arise and higher switching frequencies will be reached. This basic principle is applied in the case of the d.c regulator with an intermediate oscillating circuit, also known in short as a resonant converter. In this type of converter, the transistors are switched when they pass through the zero of an a.c. or at switch-on or switch-off [17]. In this section the configuration and operation of a d.c. regulator with a resonant circuit using 'dual' thyristors will be described [18].

9.2 Configuration

Figure 24 shows the configuration of a d.c. regulator with an intermediate oscillating circuit. As in the previous converter, one transformer \( T_1 \) powers a three-phase rectifier bridge \( I \), equipped with diodes. The input
Fig. 24  Configuration of a d.c. regulator with an intermediate oscillating circuit (resonant converter)

d.c. voltage of the monophase inverter II is obtained downstream of a smoothing filter. It comprises four dual thyristors (transistor, capacitor and diode in antiparallel). The inverter acts on a series oscillating circuit, comprising an inductance \( L \), a capacitor \( C \) and a transformer \( T_2 \). The secondary voltage of this transformer is rectified by a single-phase rectifier bridge III. As will be seen, the current \( i_d \) has high ripple. A capacitor \( C_f \) is therefore required which acts as a buffer to absorb the current ripple. The voltage \( u_d \) equal to \( u_{ch} \) is therefore practically constant.

9.3 Operation

The operation of a d.c. regulator with an intermediate oscillating circuit will be described by means of Fig. 25. By switching the two branches of the inverter alternately, an a.c. voltage \( u \) is obtained with a rectangular waveform \( +U_e \) and \( -U_e \). The current \( i \) circulating in the oscillating circuit has a practically sinusoidal waveform. If the inverter frequency \( f = 1/T \) is higher than the resonant frequency \( f_0 = \omega_0/2\pi = 1/[2\pi \sqrt{L/C}] \) of the oscillating circuit, the current \( i \) is delayed in relation to the passages through zero of the voltage \( u \).

The current \( i \) is imposed on the primary winding of transformer \( T_2 \). Taking into account the transformation ratio, the secondary current \( i_2 \) has the same waveform. The secondary current is rectified, giving a current \( i_d \). The latter therefore has a strong ripple. The ripple has to be absorbed by the buffer capacitor \( C_f \). The average value \( i_d \) is equal to the current \( i_{ch} \) in the load. It should be noted that the d.c. regulator behaves like a current source. Direct voltage \( u_d = u_{ch} \) is practically constant. The voltage \( u_1 \) at the primary of transformer \( T_2 \) is rectangular with an amplitude \( u_1' \) related to the direct voltage \( u_d \), by the transformation ratio. The rising and falling edges of \( u_1 \) coincide with the passage through zero of current \( i \).

Fig. 25  Waveforms of the various currents and voltages of a d.c. regulator with an intermediate oscillating circuit
In the intermediate oscillating circuit, relatively high frequencies can be used, from 20 kHz up to a few hundred kHz without any switching loss problems. The inverter transistors are switched on when the current passes through zero (a characteristic of the "dual" thyristor). When the current is blocked, the capacitor in parallel with the transistor limits the dU/dt which significantly reduces the blocking losses. Similarly, there are no problems for the diodes of rectifier III: they switch when the secondary current passes through zero.

9.4 Ideal characteristic

To obtain an approximation of the ideal characteristic of the d.c. regulator, the fundamental of rectangular voltages u and u1 can be used. The ohmic resistance Rch of the load appears as an equivalent resistance Re in series with the oscillating circuit L, C, as shown in Fig. 26(a).

The direct voltage \( u_d \) is given by \( u_d = I_d R_{ch} \). It is proportional to the fundamental \( 1^1 u_1 = R_e 1^1 i_1 \) of the voltage on the primary of transformer T2. On the other hand, the fundamental \( 1^1 u \) of the inverter voltage is constant. The frequency \( f \) of the inverter must be adjusted in order to regulate the fundamental wave \( 1^1 i \) of the current. As mentioned earlier, the latter is higher than the natural frequency \( f_0 \) of the oscillating circuit. It can be seen, therefore, that \( U_d/U_{d_{max}} = 1^1 u_1/1^1 u \) depends on the ratio \( f/f_0 \), from which the ideal characteristic may be deduced (Fig. 26 (b)).

![Diagram](a) ![Graph](b)

Fig. 26 Characteristic of a d.c. regulator with an intermediate oscillating circuit: (a) equivalent circuit (b) ideal characteristic.

More precise results can be obtained by keeping the waveform of voltages u and u1 rectangular and by solving the differential equations of the oscillating circuit in sections. The real characteristic is obtained by numerical simulation of the whole configuration of the d.c. regulator.

It should be noted that, in this case, the frequency \( f \) must be adjusted in order to regulate the direct voltage. This may be done very easily. All that has to be done is to adjust the blocking times of the transistors ("dual" thyristors) of the inverter.
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DC AND SLOW PULSED CONVERTER TOPOLOGIES

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ABSTRACT
Power electronics play a very important role in the supply networks of the different parts of a particle accelerator. Two of these applications are particularly interesting, namely the high-voltage supply of particle generators (Klystrons), and the low voltage and variable DC supplies for the acceleration windings as well as the particle beam deflection windings. The purpose of this paper is not to describe in detail the converters which are used for the necessary conversions, but to demonstrate how it is possible to synthesize in a systematic way the different converter structures by a thorough knowledge of the fundamental rules of power electronics and the specific requirements. This synthesis procedure should be valid for converters which fulfill the same global function but with quite distinct operating modes and performances. This procedure will be applied to the case of converters which supply the resistive or superconductor windings used for the acceleration and deflection of the particle beams. After a recall of the most fundamental rules which are necessary for the converter synthesis, we will attempt to establish a catalogue (non-exhaustive) of the different structures which can be employed in these types of applications.

1. RECALL OF FUNDAMENTAL RULES

1.1. Nature of sources

In a dipole for which the electric model contains a series inductor, the dipole current is a state variable and cannot undergo a discontinuity (there can exist no di/dt beyond a reasonable limit). Such a dipole is characterized as a current source in the power electronics terminology (Fig. 1a). The dipole current cannot have instantaneous variations because of the external circuit.

In the same manner, the duality principle indicates that any dipole with a parallel capacitor in its model should be characterized as a voltage source (Fig. 1b). The voltage across the dipole is a state variable and it cannot undergo discontinuities because of the external circuit.

![Fig. 1](image.png)
1.2. **Commutation rules. Commutation cell.**

The electronic switches of static converters are elements with commutations which allow the interconnection of impedant circuits to be modified. By considering the above definitions, these circuits act as voltage or current sources. The state variables related to these sources impose certain rules on the way these commutation are performed. Any commutation tending to impose instantaneous variations of these parameters will be prohibited. For instance, it is forbidden to turn on a switch that will connect directly two voltage sources (Fig. 2a). In the same manner, the duality principle suggests that a switch connecting directly two current sources cannot be turned off (Fig. 2b).

![Fig. 2](image)

Consequently, the interconnection between two impedant networks can be modified only if the following conditions are satisfied:

- The two networks are sources of different natures (voltage and current),
- The commutation is ensured by two switches operating in a rigorously complementary manner.

![Fig. 3](image)

The corresponding elementary circuit is that of Fig. 3 and it can present two states:

- direct connection between the voltage source and the current source (K1 is on and K2 is off). The current source imposes its current to the voltage source and inversely, the latter imposes its voltage across the current source. There will be an energy exchange between the two sources.
- In order to stop the energy flow between the two sources, we have to cancel the current in the voltage source and the voltage across the current source. This means that the two switches should be turned simultaneously into the opposite state (K1 is off and K2 is on).

The passage from one of these states to the other is done without the discontinuity of the state variables \( v \) and \( i \) of the respective sources.
The switch pair constitutes the elementary commutation cell. It is the smallest possible conversion element. In a general manner, a static converter is composed of the association of some number of current sources, voltage sources and commutation cells. Hence, each commutation involves at least one pair of switches, that is a cell. The mechanism of this commutation is described by an electric circuit, equivalent to that of Fig. 3 and by the following associated basic relationships:

\[ v_{k1} + v_{k2} = V \quad i_{k1} - i_{k2} = I \]

This relationship makes it possible to know, between the commutations, the voltage and current stresses on the switches in terms of the state variables.

During the on-state, the current passing through the switch is that of the current source. Therefore, source and switch have the same current reversibility. In the off-state, the voltage of the voltage source is applied to the switch so that it will follow the same voltage reversibility. In this way, the static characteristics of the switches can be directly deduced from the evolution of the state variables. Moreover, these two relationships allow one to study, in the switching-time scale, the commutation mechanisms of the switches (dynamic characteristics).

2. **REQUIREMENTS**

The first step of this approach consists in characterising the nature of the load and the generator. The generator is the 50 Hz 3-phase mains, which can be characterised as a 3-phase voltage source. The load is an inductor, which can include inductance and resistance in the case of a classical winding. If a superconducting inductor is used, then its resistance will be zero and the load resistance will be limited to that of the protection and contacts.

In order to precise the requirements, the following values can be taken as typical parameters and ratings:

For a resistive inductor:

\[ L = 1H, \quad R = 1\Omega, \quad t = 1s, \quad P = 40\ kW, \quad V = 200\ V, \quad I = 200\ A. \]

For a superconductor inductor, which can be considered as a pure inductance, it will therefore be supplied with a smaller voltage for the same power rating:

\[ P = 40\ kW, \quad V = 20\ V, \quad I = 2000\ A \]

In any case, the load will be characterized with no ambiguity as a current source.

The second step concerns the different operating modes that should be considered when a "current source" load is to be fed. Three fundamental modes can be distinguished:

**First or '1-quadrant' mode (Fig. 4a)**

This mode can be qualified as continuous or slowly variable (compared with the load time constant). The application of the voltage causes a current increase and, inversely, its suppression provokes the decrease of the current by dissipation in the load. The load current and voltage are permanently positive.

**Second or '2-quadrants' mode (Fig. 4b)**

This mode concerns the experimental tests, the superconductor windings and very large magnets. Here the current should be installed, maintained constant with a very good precision and then decreased very rapidly. Since the dissipation in the load of the stored energy does not allow a fast current decrease, then it should be
recovered by reversing the DC supply voltage. The load current flows in one direction but its terminal voltage is reversible.

Third 'or '4-quadrants' mode (Fig. 4c)
This operating mode concerns particularly the trajectory correction magnets which are supposed to act very rapidly in two directions. The load current is then reversible as well as its terminal voltage. The corresponding power is usually low, being of the order of 1 kW.

![Diagrams](image)

Fig. 4

3. **GENERAL STRUCTURES**

By considering the above mentioned requirements, that is a 3-phase AC voltage source at the input and a 1, 2 or 4-quadrants DC current source at the output, several general converter structures can be envisaged to satisfy to the requirements in power, rapidity, reversibility and other performance. These different general structures are illustrated in Fig. 5.
A first solution involves a direct link by means of a converter of the well-known family of rectifiers (path 1 of Fig. 5).

A second solution is obtained through an indirect link with two cascaded converters and an AC or DC intermediate stage (AC or DC link).

An AC link (with high frequency stage for ensuring the isolation and improving the performance) leads to complex structures which, up to this date, have not led to convincing industrial designs. Thus this solution will be discarded.

In the case of a DC link, we are confronted with the problem of its nature since the input and output sources are of different natures. With respect to the second converter which ensures the connection with the current source, the intermediate stage should be characterised as a voltage source and this can be achieved by the presence of capacitor C (path 2 of Fig. 5). On the contrary, with respect to the first converter, it should be characterised as a current source. It is therefore necessary to add an inductor at the first converter output, which will operate as a AC voltage-DC current converter (path 2a of Fig. 5). It is also possible to place this inductor at the AC side of the converter, in which case the nature of the AC source will be modified. Then the first converter will be of AC current-DC voltage type (path 2b of Fig. 5). All these possibilities will be discussed in the next section.

To obtain the good performance required in these applications, it may also be necessary to provide passive filters and even in some cases active filters at the load terminal.

4. **STUDY OF DIFFERENT STRUCTURES**

4.1 **Direct Converters: Rectifiers**

Since the voltage source is an AC one, the switches should be voltage reversible. If the current source is not alternating, then the switches are one-directional in current. The static characteristics of the thyristor are naturally adapted to this condition (Fig. 6).

![Figure 6](image_url)

Figure 7 represents the well-known schematic diagram of the corresponding structure (six-pulse three phase-bridge connection). Voltage and current waveforms at the DC side are also illustrated in this figure, where the need for filters to eliminate the residual ripple is evident.

For high power levels, the ripple can be reduced by an increase in pulse number. For example, phase-shift bridge arrangements in series or parallel configurations (Figs. 8 and 9) can raise the pulse number from 6 to 12.
The phase shift is achieved by employing an input transformer with two secondaries, one star-connected and the other delta-connected. Some complex configurations can increase the pulse number up to 24 or even 48.

These converters are naturally voltage-reversible and therefore allow 1- and 2-quadrants operating modes. Half-controlled configurations (diodes and thyristors), or the use of a free-wheeling diode, allow only the 1-quadrant operating mode. If a 4-quadrants operation is required, then the current source and also the switches should be current-reversible. Such an operating mode is achieved by using inverse-parallel thyristors (converters of cyclo-converter type, Fig. 10) or even inverse-parallel converters (crossed bridges with or without circulating current).

![Diagram of inverse parallel thyristors or converters](image)

**Fig. 10**

This type of converter can then operate in the three operating modes mentioned above but the following facts should be pointed out:

- The three-phase transformer operates at the industrial frequency (50 Hz),
- The response times are quite large,
- The current ripple is much larger than the minimum tolerated level,
- The increase of the pulse number from 6 to 48 for reducing the ripple magnitude involves more complex control strategies and increases the instability risks,
- The necessary passive or active filters have also complex structures.

4.2 **Indirect converter with DC link**

This conversion scheme is realized by means of two converters and an intermediate stage which behaves as a DC voltage source (capacitor C) with respect to the second converter. The latter will then be, in all cases, of DC voltage-DC current or DC voltage-AC current type (according to the reversibility requirements). It has already been mentioned that in order to respect the opposite nature of the sources (the nature of the input source should be the opposite of the output one) an inductor L should be inserted between the first converter output terminals and the voltage source, or placed at the input of the first converter. Several structures are possible for the latter.

4.2.1. **Input converter**

*First structure of input converter*

The inductor is placed at the DC side of the first converter. The corresponding schematic diagram is shown in Fig. 11 where the first AC voltage-DC current converter is followed by the second one of DC voltage-DC(or
AC) current type. The first converter then belongs to the family of conventional thyristorised rectifiers. Even though this configuration respects the interconnection rules between sources, it does not allow reversibility. It can be noted that the DC-current output of the first converter is only voltage-reversible and that the DC-voltage input of the second converter is only current-reversible. Then, although these two converters are naturally reversible if they are separated, there will exist a problem of intermediate-stage mismatch, as far as the reversibility principle is concerned.

Fig. 11

If one wishes to ensure the possibility of reversible operation and maintain, at the same time, the choice of no voltage-reversible DC voltage source as the intermediate stage, then its current reversibility should be guaranteed. In order to achieve this, the input converter should be current reversible on the DC side. As mentioned earlier in the case of a direct converter, this can be done by means of two inverse-parallel converters. Figures 12 and 13 represent the operation principle and the realisation of this network, respectively.

Fig. 12

Fig. 13
In the case where it is not worthwhile recovering the energy in the reversible operations, then one solution consists in its dissipation in the intermediate stage. During the direct interval, the energy is provided by the top rectifier (positive current flow) and in the opposite direction (negative current flow), the electrical energy is dissipated in a resistance placed in parallel with the capacitor (Fig. 14). The current starts to flow in this resistance from the moment the voltage across the capacitor exceeds a given reference value. Since only the second converter is reversible, then the first converter can be a simple diode rectifier.

![Figure 14](image)

*Second structure of input converter*

In this case the inductor is placed on the AC side of the first converter (Fig. 15). The intermediate stage will then behave as a DC voltage source with respect to both the first converter output and the input of the second. Here the first converter is of AC current-DC voltage type while the second one is of DC voltage-DC(or AC) current type. The current reversibility is therefore natural in the intermediate stage and, consequently, conventional structures can be directly used.

![Figure 15](image)

![Figure 16](image)
The first converter can be a PWM voltage inverter, but operating in rectifier mode (Fig. 16). The advantage of this technique is that the AC mains current can be controlled by the modulating waveform. In this way its phase is controlled and its harmonic content can be reduced. Figure 17 illustrates these possibilities:

17a – comparison between current waveforms of a diode-bridge rectifier and a PWM bridge
17b – possibility of a zero phase-shift between the voltage and current
17c – the current lags the voltage
17d – the current leads the voltage.

![Diagrams](image)

Fig. 17

4.2.2. Output converter

This is a DC voltage - DC current converter which can be current reversible or not. Its structure depends on an additional parameter, specified by the overall requirements, namely the ohmic isolation. This can be achieved by means of a low- or high-frequency transformer inserted in an AC intermediate stage.

*Low-frequency ohmic isolation*

If the ohmic isolation is imposed, as is usual, it can be ensured by a 50Hz low-frequency transformer in the input network of the first converter. In this case the second conversion is direct. The corresponding structures are well-known and depend on the required reversibility mode.

1-quadrant mode: the voltage and current quantities are not reversible so that both switches are also non-reversible in voltage and current. This leads to the classical chopper scheme of Fig. 18a.
2-quadrants mode: the current source is only voltage-reversible, leading to the bridge scheme of Fig. 18b in which the switches are non-reversible in voltage and current. This is a voltage-reversible chopper.

4-quadrants mode: the current source is voltage and current reversible suggesting a bridge scheme with current-reversible and voltage-non reversible switches. This gives a voltage and current reversible chopper (Fig 18c).

High frequency ohmic isolation

The use of a HF transformer allows, besides the isolation possibility, the response time and the current ripple to be reduced, even with low-frequency-ratio structures in the input.

Direct DC-AC-DC converters

The conversion is obtained by the direct connection of a DC-HF AC converter and an HF AC-DC converter. The link circuit between the two converters is the HF transformer to guarantee the ohmic isolation. There is no intermediate stage with an auxiliary source as in the case of indirect converters, since the input source is a voltage source and the output source is a current one. For the first converter the input of the second converter behaves as a current source through the transformer. Similarly, the output of the first converter is a voltage source for the second converter. According to the desired operating mode, the different structures of Figs. 19–21 can be used:
Figure 19. 1-quadrant converter. This is a non-reversible scheme of the FORWARD type. The first converter presents a semi-reversibility in current at the DC side in order to allow the flow of the magnetising current of the transformer (the input capacitor suffices to ensure the corresponding energy balance). Thus an asymmetrical bridge has to be used and a simple diode rectifier is placed in the transformer secondary.

Figure 20. 2-quadrants converter. This involves an intermediate AC voltage stage. An inverter produces an AC voltage that is used to feed the rectifier input through the transformer. The rectifier output is a DC current source being voltage but not current-reversible. This leads to a bridge with four switches which are voltage-reversible but non-reversible in current. The first converter, with a DC voltage source at the input, encounters an AC current source at its output, created by the rectifier. Here the necessary circuits are a bridge with four current reversible and voltage-non-reversible switches as well as a conventional inverter structure. The transformer secondary behaves with respect to the rectifier as an AC voltage generator, with an AC current having a phase-shift with the voltage. If the current lags the voltage, then the rectifier switches will be turn-on controlled, thyristors constituting the most interesting solution. The inverter sees, therefore, the transformer as an AC current receiver with a voltage lagging the current. Here the inverter switches are turn-off controlled and consequently both converters are of dual type.

Figure 21. 4-quadrants converter. The second converter should be voltage-reversible at the AC side and current-reversible at the DC side. The switches have therefore to be voltage and current reversible, which is achieved by means of the inverse-parallel connection of the switches and a structure of the cyclo-converter.
type. The first converter, being of the voltage inverter type as in the preceding case, already possesses the required reversibility. If the rectifier is constituted by turn-on controlled switches (inverse-parallel thyristors) then the inverter should have turn-off controlled switches.

Fig. 21

*Indirect DC-AC-DC converter*

If a series inductor is placed in the primary or secondary circuit of the HF link transformer, the conversion becomes indirect. This means that the first conversion should be of DC voltage-AC current type while the second is of AC current-DC voltage type. It can then be observed that there exists an incompatibility concerning the nature of the input and output sources of the second converter. The solution consists in placing a parallel capacitor with the output current source. In this way the output circuit behaves as a DC voltage source. It can be noted that if the load is preceded by a passive L-C filter (this filter could be supposed, up to this point, to be included in the equivalent current source of the load network) the same result can be obtained if the filtering inductor is rather placed at the AC input of the second converter instead the DC output. The L-C filter can therefore be considered as being part of the converter.

In 1-quadrant operating mode, the first converter is a voltage inverter and the second converter can be a simple diode rectifier (Fig. 22). The latter is placed between an AC current source in the input and a DC voltage source in the output so that its input voltage will be rectangular. The differential AC voltage between the rectangular voltage at the inverter output and the one at the rectifier input will appear across the inductor L, which supports an AC current. Since a high frequency is present, it is not possible to think of a PWM operation to ensure the power control. But, the inductor L placed in the HF network constitutes a first-order filter so that each

Fig. 22
increase in the operating frequency imposed by the inverter is accompanied by a decrease in the transmitted power. Consequently, this system requires large frequency dynamics in order to achieve large power dynamics.

In order to reduce this frequency dynamic, a second-order filter can be used by a series L-C circuit as shown in Fig. 23 (the capacitor C eliminates at the same time an eventual DC current component in the transformer). The converter is therefore of the series resonance type (SRC).

![Fig. 23](image)

The power reaches its maximum value when the operating frequency is equal to the natural frequency $f_0$ of the series resonant circuit. The power regulation can therefore be performed at a frequency below or above $f_0$.

The switching mechanism of the inverter depends on the retained solution. For $f < f_0$, the switches should be turn-on controlled while they are turn-off controlled for $f > f_0$.

In order to achieve a further increase in the operating frequency range, a third-order filter can be employed (Fig. 24).

![Fig. 24](image)

Finally, a last solution in the 1 quadrant operating mode consists in bringing the L-C filter to the primary of the HF transformer. Then the intermediate stage behaves as an AC current source with respect to the first converter and as an AC voltage source with respect to the second. The AC voltage is rectified by the diode bridge and then applied directly to the DC source (Fig. 25).
In the 2-quadrants mode, the current reversibility of the output converter should be ensured, since the input converter is naturally reversible, for instance of the series resonance type, similar to that of Fig. 23. An AC current-DC voltage rectifier can therefore be used, with a structure similar to that of a DC voltage-AC current inverter (Fig. 26).

For a 4-quadrants operation it is necessary to ensure the voltage reversibility at the DC source terminals. Similarly to the two inverse-parallel bridges of Fig. 14, in order to guarantee the DC current reversibility at the output of a voltage-current converter, here the DC voltage reversibility can be ensured by two identical bridges connected in inverse-series (Fig. 27). The voltage appearing across the current source is the differential voltage $V_A - V_B$ of the two bridges $A$ and $B$. Therefore, at each instant it takes the sign of the greater voltage. If the average value of the two voltages is the same, then the voltage across the current source is alternating. Each of these two bridges is an AC current-DC voltage rectifier and consequently their switches are current reversible but non-reversible in voltage.
5. **CONCLUSION**

In conclusion, we can consider the synthesis of all the different configurations studied above, in order to distinguish those which present particular interests for the considered applications.

In the case of direct conversion and concerning the input converter in the case of indirect conversion, three AC-DC voltage reversible structures can be retained:

a – The association of inverse-parallel thyristors or crossed three-phase bridges (Fig. 13)
b – The three-phase rectifier bridge with a dissipation resistance (Fig. 14)
c – The PWM inverter-rectifier (Fig. 16)

In the case of indirect conversion, the most utilised structures for the output DC-DC converter are:

1-quadrant mode
a – Direct DC-DC converter: the BUCK scheme (Fig. 18)
b – Voltage AC link and HF transformer: the FORWARD scheme (Fig. 19)
c – Current AC link and HF transformer: SRC scheme (Fig. 23)

2-quadrants mode
– Cascade of a voltage inverter and voltage rectifier, with dual commutations (Fig. 20)

4-quadrants mode
a – Cascade of an inverter and a cyclo-converter with HF AC-voltage link (Fig. 21)
b – Resonant converter with differential voltage output and HF AC-current link (Fig. 27)
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ACHIEVING HIGH PERFORMANCE

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Abstract
The machines of the high energy physics world require high performance power converters. Correctly, one immediately thinks of high precision, stability and dynamic performance and this presentation covers these topics which are also taken up more fully in other specific presentations. However, other performance criteria such as reliability, cost and environmental impact are also considered. Since the control loops and reference sources are treated elsewhere, the presentation analyses two types of gate firing circuits and ends with a practical example of a power converter design for a collider.

1. INTRODUCTION

This very course is about achieving high performance in power converters. Each and every presentation touches on this subject. The role of this presentation is to highlight some of these points and show how each subject is cemented into a whole to achieve overall high performance.

High performance might make one immediately think of very precise or fast amps and volts, much as high performance in the car world brings to mind top speed and acceleration. These are important but by no means the only criteria for high performance. There are many other factors which contribute to high performance. Consider for a moment the following:

i) reliability;
ii) repairability, availability;
iii) purchase and running cost;
iv) effect on environment.

Achieving high performance is a compromise. Never let the blind pursuit of one factor in high performance totally wreck another. Each factor should be tuned to the optimum so that each goal is achieved. It is useless to have a $10^{-6}$ stability with high dynamic performance if the power converter is always breaking down and is too complicated for anyone to repair. This is even more ridiculous if $10^{-4}$ stability would have done and only slow ramps are used. If on top of that the power converter costs a fortune to buy from a very specialized company and consumes spare parts and electricity at a high rate, it may not be considered high performant by a project leader or director. It might also pull very high peak currents from the supply network while generating large harmonics with a poor power factor. The electricity service would not consider that a high performance power converter. It might also overheat buildings, make a terrible noise, use up a lot of building space and be completely user unfriendly in every respect!

Before going on any further, let us first consider getting the right performance specifications from the beginning.
2. GETTING YOUR HIGH PERFORMANCE SPECIFICATION

It has been shown [1] that the power converters are a vital performance element in any accelerator. The precise translation of the beam's and/or the machine's needs into amps, volts and their respective tolerances is a vital part of achieving high performance. Get this wrong and your compromises may be wrongly weighted. It is also important to try and understand the exact operating modes that might be used in terms of cycles, sequences etc.

Never accept a tolerance or requirement at face value. Always find out where it comes from and if it is really necessary. Safety factors get added to safety factors and before you know where you are, a very very tight tolerance appears which may not be necessary. On the other hand, never sail too close to the wind only just achieving a performance criteria if it is really vital to the machine.

Find out about your loads. Are they really just L and R or perhaps something containing many more parasitic elements which add up to a complex circuit? What is the real transfer function from the power converter terminals to the parameter controlling the beam? What is the effect of shielding or the vacuum chamber on ripple and acceleration?

Will the communication network and timing system be sufficient for the type of performance you have to achieve? How will the machine be operated and what diagnostics will be needed in the control room? Has anyone even thought about it?

Discuss in detail your power requirements from the network. Optimize the voltage levels, power factor correction and harmonic filters to achieve a stable system. Find out about any special security features that may be necessary and interlocking procedures. A discussion with the civil engineering branch to find out the cost of space, cooling and installations will help optimize your solutions.

Having done this, it should be possible to avoid any false economies and convince your direction that an optimum design has been achieved. Further it should be clear that, while power converters are relatively cheap in the machine context, their importance is vital.

The whole process may take several iterations, but it is well worth the effort if by the process you can achieve high performance in several domains.

Now let us consider some of those more mundane performance factors we mentioned earlier.

3. MUNDANE PERFORMANCE FACTORS

3.1 Reliability

This is a factor too often neglected by those in pursuit of pure high performance. To my mind, it is the number one performance goal. Simplicity is the name of the game. Excessive component counts must diminish reliability. Therefore, design up to the performance required using the least number of components and stop
there. Never add a component unless it is needed. If a transformer and a diode fulfil the performance needs, don’t put in a fully controlled thyristor set with an active filter just because it looks good in the design report. However, if higher performance can be achieved for nothing by a judicious choice of a type of topology or operating system then take that. A simple design offering high performance is the clever one, not the complex overkill.

Avoid electro-mechanical components wherever possible. Strive to achieve a real static converter. Circuit-breakers are normally required as the ultimate protection element and therefore cannot always be avoided. (Other switches may prove more reliable.) However, make them easy to repair or better still to replace. Limit the number of operating cycles. Never turn power converters off unless really necessary. Better set them to a low-level output setting and leave the circuit-breaker closed. Most failures occur at switch on either because the breaker does not close, or because the electrical and thermal shocks produced by the procedure kills components. Try and arrange that the operation of the machine minimizes switch-offs. Fans should also be avoided if possible as well as tap-changers and roller regulators. Routine maintenance, although costly, can help considerably the reliability of these items.

Always try and achieve reasonably low temperatures. High temperature kills insulation and dielectrics. Yes, you can run a piece of copper at high temperatures but consider the components near to it. Failures, particularly in electronics, seem to increase dramatically once the ambient temperature goes above 30°C. Here a judiciously placed fan may actually increase reliability, if it is off the critical reliability path.

The critical reliability path can be defined as those components whose failure will stop the converter working. These types of components should be minimized and conservatively rated. Once again temperature and voltage are the main killers. On the other hand, some additional auxiliary circuits may increase reliability while not contributing to it (make sure their failure does not reduce reliability). Diagnostic circuits will greatly help repair and preventive maintenance but make sure they do not cause failures themselves.

Protection circuits have the hardest role and are covered elsewhere [2]. Their design must be fail-safe, always working when required but never tripping when not. Choose very carefully your protection circuits and never include an interlock unless it is really necessary. Perhaps spending a little more money on a more robust component may be more cost effective. Again, simplicity in protection circuits is normally the key to reliable operation.

3.2 Repairability, Availability

If, despite all your efforts, a failure does occur, (undoubtedly because someone did something wrong) then repairability is all important. It may be blatantly obvious to you, as the equipment designer, what has gone wrong and what to do about it, but think of the poor chap at 3.00 a.m. who has been on shift for 12 hours. The higher the level of technology then the more modular the system needs to be so that simple replacements can be made. Don’t expect anyone to repair a 40 kHz, 30 kW inverter or a microprocessor card in the early hours of the morning with several angry physicists waiting for beam.
If possible a "universal" spare converter might well be justified. This can keep repair times (from the machine point of view) low and allow non-specialized staff to make the replacement.

Clear diagnostics are vital both locally and more importantly at distance, if the machine is large. In this way, interventions can be planned from the control room. This point is covered in Ref. [3].

While high reliability must be the ultimate goal, do not forget that actual repair time is important if maximum availability is to be achieved, and efficiency of the machine in terms of availability to the physicist is what counts. Depending on fill times etc., two failures needing 10 minutes to repair might be better than one needing a day. Total time lost is the factor in judging performance in this respect.

A high performance power converter is one that is already "run-in" when handed over to the machine operators. It should be fully available from the beginning. This means that power converters should undergo very thorough type and routine tests and have undergone several hours of operation on test stands or eventually during cold check-outs of the machine. Once beams arrive the operations crew don't appreciate power converter teething problems.

3.3 Purchase and Running Costs

The "any price for performance" that one sees in the car world cannot be applied to power converters for accelerators. Our directors believe that performance should be achieved at a reasonable price, and they are right. However, whatever you do, do not fall into the trap of false economies. Upgrades and modifications at a later date can be financially disastrous. Fight against anything that might compromise the successful procurement of the power converters and their final performance. These points will be covered in Ref. [4].

As we should not forget the overall system performance, neither should we forget the system costs. In choosing the correct topology to meet our own performance and cost criteria, we should not forget such things as building, cooling and cabling costs.

With the ever increasing demands to cut staff to a minimum, our designs should take this into consideration. We have already considered high reliability to cut interventions to a minimum but do not forget the initial costs of complex installations and test procedure as well as that of staff training.

Likewise for running costs, the converter should have high conversion efficiency, low spares needs and consumption and a minimum need in routine maintenance.

3.4 Environmental Performance

Again, making the comparison with the car industry, we have to be aware of the effect of our power converters on other systems and people. Audible noise is an obvious one but electrical noise may be more important. It is no longer acceptable to generate large quantities of RFI or EMI and expect people to live with it. In fact, it could completely close down an experiment's detector. Power converters should wherever possible be
housed in their own building away from any sensitive equipment and connected to their own network. It has to be admitted that they are electrically noisy so don’t tempt fate.

The power converters’ own electronics have to live with the noise however. Precisions in the ppm region cannot be achieved by locating DAC, ADC, etc away from the power converters. Drive and protection electronics likewise need to ‘live’ in the power converters’ environment. Since there are limits to just how much can be done to reduce noise in the power converter, the electronics have to be made less susceptible to it. Careful control of currents (particularly noise) on the 0 V rails and the earthing of one critical star point are important. Galvanic isolation for digital signals is always helpful, as well as the use of differential measurement in the analogue world. Strive to achieve symmetry in the power circuit, feedback and control signals.

There are some instances when it is clear that electronics noise will be created. During switch-on of a large transformer, or during commutation of a semiconductor are some of the cases. It is often known at what moment these will occur and sensitive circuits can be de-sensitized or blocked during these moments. Driver or firing circuits are often blocked for some microseconds following the command for commutation. In this way, the generation of further spurious signals can be avoided. Likewise many circuits (but not normally protection circuits) can be blocked during switch-on until all transients have died down.

Power converter designers must not only consider the load, but also the network supplying the converter. Low mains harmonic distortion should be a goal by limiting the generation of non-characteristic currents. In general the supply companies want us to consume a sinusoidal current at near unity power factor. Certain topologies are far from this. Take the typical small SMPS with its diode and capacitor input stage. Acceptable in small quantities, but a disaster when you have several hundreds together. The aim should be towards 3-phase or at least input stages with LC filters taking current over the longest possible conduction period (3-phase also helps with single-phase mains dips).

Topologies having naturally the lowest harmonic distortions should be chosen wherever feasible. Phase shifting of larger units can allow harmonic currents to be exchanged between them but if they are phase-controlled then this advantage can be lost at different operating points. Most attempts to produce the ideal input stage tend to be very complex and are not always cost-effective nor reliable. They can also become unstable depending on the number of units connected or not to the network. A better technical solution at a lower cost can often be achieved by a "clean-up" at the system level. A well designed set of harmonic filters and reactive compensators at large consumption points can achieve this while assuring mains network stability for all operating conditions.

4 PRECISION

Now assuming we have a superbly reliable, minimum cost, low pollution power converter, what should it do for the machine? No matter what type of machine, the power converter needs to respond to a reference command as precisely as it can, not deviating dynamically nor statically from that requirement. To do this, the power converter has the following essential elements. They are shown in figure 1.
Fig. 1 Generalized block diagram of the precision components of a power converter for a magnet load

DAC - Digital to analogue converter (assuming that it is an analogue regulation system) to give the reference;
Transducers - To feed back analogue likenesses of the parameter or parameters to be regulated;
Control loops - To stabilize those parameters;
Modulators - To cause the changes in the parameters as dictated by the control loops;
Filters - To reduce unwanted noise.

Precision can be split into three distinct parts as shown in figure 2.
i) ripple;
ii) short term;
iii) long term.

Ripple is determined by the modulators and the output filters. For a magnet power converter the magnet and the vacuum chamber also constitute the output filter. The amplitude and frequency of the modulator, be it switch-mode or thyristor phase control, will dictate the constant and continuous perturbation caused to the beam by the ripple. Since it is continuous, it is normally defined to be very small (say 10^-6). Higher frequency modulators producing low peak-to-peak amplitudes need less passive (or active) filtering.

The modulators and filters are also important in the rejection of mains perturbations which come under the heading of short term precision. Again, the faster the modulator the smaller the filter can be. For D.C. systems, a large passive filter may be advantageous but in a pulsed system would hinder dynamic response. In colliders the power converters' ability to reject mains disturbances is paramount. A lost beam due to a mains dip can mean many lost hours, which is not the case for cyclic machines. The control loops have to achieve the maximum bandwidth with a reasonably damped response. If not, the tune of the machine may be altered or the aperture limit reached and the beam lost. Alternatively, if there is a continuous mains variation, because of a pulsing load such as an injector, then its effect may be seen on the beam with resultant bad physics. These effects are summed up in figure 3, which shows the rejection curves for typical switch-mode and thyristor power converters. In general mains rejection should achieve a 10^-5 stability. Hence a 1% mains change would need a 10^3 rejection at the relevant frequency and a 10% mains change would need 10^4 rejection.
**Fig. 2** The components of precision

**Fig. 3** Optimistic rejection curves for power converters
The DAC, Main Current Transducer and comparator amplifier determine the long-term precision and reproducibility of the power converter since their bandwidth is normally limited to a few hertz. They are an important factor in the reproducibility of the machine from run to run, from operating period to operating period and of the general drift of the machine. While a certain amount of tuning can be done via the beam feedback, and eventually magnet field feedback, it is the current reproducibility that is important. More and more today high-precision DCCTs (direct current current transformers) perform the role of the main current transducer. They offer galvanic isolation from the power circuit and high signal levels both of which contribute to low noise. The zero-flux type of current transducer (Fig. 4) has now become the norm. By detecting in one way or the other the absence of field in its core when injecting a balancing ampere-turns, the primary current can be determined to resolutions in the region of $10^{-7}$. The problem comes when converting the secondary current (normally in the region of 100 mA to 5 A) into a voltage for comparison with the reference voltage. Various alloys now exist (nickel-chrome etc) to make up burden resistors which can have very flat temperature coefficients and low self heating. Ageing effects have also been minimized. Depending on the power and stability required, the burden resistors can be temperature-stabilized and the voltage amplified for comparison. Today burdens having total tolerances in the region of 50 ppm can be attained while below 20 ppm is possible with extra care. These types of DCCT's have a bandwidth up to 10 kHz and noise levels in the microvolt region which means that they can be used in fairly fast applications and are rarely limiting factors to performance.

![Schematic of a zero-flux DCCT](image)

Fig. 4 Schematic of a zero-flux DCCT

The digital to analogue converter is covered in another presentation. It has often proved to be the most critical element. The communications and control systems being digital while the control loops remain analogue means that the DAC is still with us. If digital DCCT's could be produced having the same speed and accuracy as their analogue equivalent, then the way would be open for adaptive digital control loops and the elimination of the DAC completely.

Today the modern operational amplifier has sufficient gain, low-offset and drift that it is fairly straightforward, with one or two precautions, to obtain a very precise comparator which can almost be forgotten. Likewise capacitors with low leakage are available such that an integrator with very high D.C. gain can be obtained. In general, we have tried to use voltage comparison and gain one buffer (Fig. 5) so as to eliminate any further precision resistors. In this way the D.C. performance of the power converter depends solely on the DAC and DCCT. A minimum of adjustable potentiometers are also used so as to minimize screwdriver twisting!
5. OTHER PERFORMANCE ELEMENTS

The control loops are quite clearly a very important element in achieving high performance [5]. Likewise, the transducers will also be covered by a separate presentation.

The missing part, at least for a thyristor phase-controlled power converter is the firing circuit which performs the reconversion of the analogue control loop signal back to the digital thyristor pulses. Again, this seems to cry out for a digital control.

Its correct reliable operation is fundamental to the well being of line-commutated thyristor converters. It is often the most complex circuit in terms of component count in the regulation of the converter, and if incorrectly designed can be the source of considerable degradation of performance and reliability.

A firing circuit should have as many of the following features as possible:

i) simplicity, low component count;
ii) ease of adjustment or preferably none;
iii) high bandwidth and good linearity to facilitate control loops;
iv) no sub-harmonic generation in the D.C. output;
v) efficient, autonomous end-stops;
vi) high network noise immunity and lack of system cross-talk;
vii) rapid recovery from transient conditions;
viii) no amplification of converter or network asymmetry;
ix) ease of application and understanding by technical staff.

The LEP-PC group has developed two types of firing circuits. The first, operating on principles established in the ISR and SPS machines, is known as the Cassel (Ainsworth) Van der Meer circuit. The second, operating on completely different principles is known arbitrarily as the ASAD circuit. Both are actually lodged on the same card, since they use common pulse distribution and end-stop circuits. One can choose by switch position which is to be used.
5.1 Operating Principle of the Cassel/Ainsworth/Van der Meer Circuit

The interesting feature of this circuit is that it uses information contained in the output D.C. waveform in order to generate firing pulses. As such, it operates in the closed-loop mode and forms an integral part of the converter function. This elegant feature can in a minority of circumstances and applications be a disadvantage, but in general it fulfills most of the previously mentioned criteria.

Fig. 6 Operating principle/performance of Cassel/Ainsworth circuit
The theory of operation is shown in figure 6a, where the mean value of the output D.C. waveform A is compared with the desired reference level B, using an integrator. Since the output waveform of the thyristor bridge is made up of several parts of different sine waves of the 3-phase network which, with the exception of near maximum and minimum operation, can be considered as a saw-tooth waveform, then the integrator output C will be parabolic and cross the fixed switching reference once the positive and negative areas of the output waveform A are equal. At this moment a firing pulse is generated. Should a disturbance occur, as shown in figure 6b, then the integrator will again assure that both positive and negative portions are equal and hence will generate a delayed firing pulse. The resultant portion of positive D.C. output is smaller, and once again will be compensated by the integrator. From then on a permanent oscillation sets in.

In figure 6c, the zero switching reference has been replaced by a saw-tooth such that its slope equals the initial slope of the integrator's parabolic output. In the case of a disturbance, as before, the firing pulse is delayed to a lesser extent such that the parabolic falls back onto the value it would have had with no disturbance. Hence, correction occurs in the shortest possible time and the system is now stable.

As is often the case, practice is not always as straightforward as theory. The output waveform is made up of sine waves and hence the slope of the saw-tooth should be varied for different output levels and in particular for operation with discontinuous conduction. The phenomenon of overlap has not been taken into account and this has the tendency to slow down the action of the circuit. The exact amplitude of the saw-tooth is a compromise between speed of response and stability.

In general, these shortcomings can either be overcome by increased circuit complexity or have a minor effect on overall performance. This circuit, or various derivatives of it, is used in about half the LEP magnet converters and all of the SPS power converters.

5.2 Alternative Open-loop Firing Circuit (ASAD)

In certain applications, notably for primary control, high-current converters, complex loop situations and problems of switch-in, a fast open-loop circuit can give better performance. With this in mind a new circuit was developed.

The new circuit had to possess as many of the afore-mentioned features as possible, while working in open-loop mode. It had also to possess features making it a suitable candidate for the high-power klystron converters of LEP. In particular, it had to have the very highest dynamic response so that control loops, and therefore performance of the converters, could be optimized.

The ASAD circuit, instead of observing the integral of the sine wave currently appearing in the output, adopts a different point of view. It works on the philosophy that once a thyristor has been fired, and therefore committed, it is better to observe the state of the next sine wave. Hence, once a thyristor has been fired, a solid-state selector looks at the next sine wave C which will be applied to the output and compares this with the input reference B (Fig. 7a). At the crossing point a new thyristor is fired, producing waveform D, and the circuit immediately moves onto the next sine wave. In this way, the circuit is always ahead of what is produced at the output and can thus respond without delay to any change of the input reference.
Fig. 7a. Simplified operating principle of ASAD circuit.
In practice the selector does not look directly at the 3-phase input, but instead at an integrated version of this. The integrator causes a phase shift of 90° at 50 Hz (Fig. 7b) thus presenting a cosine function for $0^\circ < \alpha < 180^\circ$ ($\alpha$ being the firing delay). Because of this, operation up to maximum ($\alpha = 0^\circ$) can now be achieved, and the transfer function of voltage reference to bridge output is now linearized since both follow the well known cosine law. The last and very important role of the integrator is to provide a high level of noise immunity so as to avoid the normal problems associated with zero-crossing detection.

### 5.2.1 Linearity

As expected, the cosine function of the thyristor bridge has been corrected. Photo 1a shows the smoothed D.C. output following exactly the triangular input reference from 0 to 100%. Photo 1b shows the effect of non-continuous conduction where a non-linearity can be seen during the start of the rising edge. The rapid desaturation from 100% can also be observed.

![Graph showing continuous conduction](image)

100 %

0 %

**a) Continuous conduction**

- Smoothed D.C. output
- $T = 200$ ms/division
- Input reference

![Graph showing non-continuous conduction](image)

**b) Non continuous conduction**

- Smoothed D.C. output
- $T = 200$ ms/division
- Input reference

Photo 1 ASAD Linearity

### 5.2.2 Dynamic response

According to theory, this firing circuit should be extremely rapid in its responses. Photo 2 shows the circuit responding to various waveforms and in particular the often delicate passage from rectification to inversion. For comparison, the response of a Cassel/Ainsworth/Van der Meer circuit to a step input is also shown.
5.2.3 Frequency response

The frequency response of the circuit was measured using a Transfer Function Analyser (TFA). Generally accepted theory suggested that the best frequency response one could expect, bearing in mind the sampled nature of the thyristor bridge, would be a 3 dB down point at 150 Hz with absolutely no response beyond 300 Hz.
The ASAD circuit seemed to demonstrate a complete disrespect for this theory and showed a relatively flat response up to about 700 Hz, rolling off beyond that point. Figure 8 shows this response which is compared with that of the Cassel/Ainsworth/Van der Meer for various amplitudes of saw tooth correction. The reason for this apparent excellent frequency response can be seen in photo 3, where the output waveform is shown for various input sinusoidal waveforms. The 300 Hz fundamental waveform of the thyristor bridge is clearly modulated even above 300 Hz and that is why sampling theory should not be applied. Why does it roll off at all then? With the ASAD circuit the frequency response of a thyristor bridge is not governed by its fundamental supply frequency but by the delay in time from the moment the circuit asks for a thyristor to be fired and the completion of the commutation. In other words, it's governed by the overlap angle which of course depends on the supply inductance and the load current.

![ASAD following 40 Hz sinusoidal input signal](image)

![ASAD following 200 Hz sinusoidal input signal](image)

![ASAD following 400 Hz sinusoidal input signal](image)

Photo 3 Frequency response of ASAD
5.2.4 **Ability to work with high gain loops**

The high frequency response, coupled with the good linearity, would suggest that this circuit would be ideal for integration into high performance control loops, which indeed has proved to be the case.

5.2.5 **Low level operation**

As would be expected, operation at very low level is now possible. The exact levels depend on the type of loop and clamp circuits used to feed the firing circuit.

The ASAD circuit has been employed in two types of converter for LEP. The first is the 100 kV/40 A klystron converter, where the circuit is used to fire a 6-phase thyristor line-controller (~5 MVA) used in a primary control configuration. In this application its open-loop nature was needed as well as its very accurate generation of pulses at 180° separation which thus assures very little D.C. content in the primary of the load transformer.

The second application required the very high speed of the ASAD circuit. A dual-converter is used for some of the bipolar converters of LEP, where two thyristor bridges are connected in anti-parallel and a current circulated between them. The loop which controls this current circulation must be as fast as possible so as to avoid excessive overload currents and allow the subsequent output loops to have, themselves, a high bandwidth.
6. **HOW DO WE DESIGN A SIMPLE HIGH-PERFORMANCE THYRISTOR CONVERTER**

Since switch-mode is discussed in reference [6], I would like to run through the design of the power part of a thyristor power converter for collider application [7].

The specification for the machine is:
- I nominal = 300 A
- U nominal = 300 V
- Magnet L = 1 H (vacuum chamber time constant = 5 ms)
- Ripple current = 5 x 10^{-6} of max
- Mains variation = ±10%
- Precision/Reproducibility = 10^{-4}
- Range = 20 : 1
- Temperature variation = 10 - 30°C.

![Diagram of a thyristor power converter](image)

**Fig. 9** Basic elements of a 6-pulse thyristor power converter

6.1 **DCCT**

A 300 A/10 V DCCT of the zero-flux type will do the job. It should have an overall performance better than 50 ppm. Therefore less than 0.5 ppm/°C, practically zero voltage coefficient and a drift better than 20 ppm/six months. Self heating should only add 10 or 20 ppm. It must be the last element in the power converter.

6.2 **Free-wheel diode (FWD)**

Assuming that the power converter does not have to operate in inversion in order to force the discharge of the magnet then a free-wheel diode will be used either at the output or across the bridge.

a) **On the output** - To protect any electrolytic capacitors and discharge the magnet when a rapid trip occurs. It must have adequate thermal mass to absorb the decay current in the case of total coolant loss due to a mains failure. This is normally only a problem for superconducting magnets.
b) **On the bridge** - In this role the diode will not only have to discharge the magnet but will commutate off the thyristors at zero voltage, thus stopping inversion. The current flowing through the diode will depend on the firing angle and the load mismatch (Fig. 10).

![Diagram of FWD current values and power losses for the LEP main ring converter](image)

Fig. 10 FWD current values and power losses for the LEP main ring converter

6.3 **Filtering**

The magnet time constant is \[ \frac{L}{R} = \frac{1 \times 300}{300} = 1 \text{ s}. \] Break point \[ \frac{1}{2\pi} = 0.16 \text{ Hz}. \]

Input ripple voltage (peak to peak) equals:

\[ \sqrt{2} U_{\text{line}} \]

Assuming a full-wave 3-phase bridge:

\[ U_{\text{line}} = \frac{\pi}{3\sqrt{2}} U_{\text{d.c.}} = \frac{U_{\text{d.c.}}}{1.35} \text{ at } \alpha = 0^\circ \]

\[ = \frac{300}{1.35} = 222 \text{ volts.} \]
A factor of about 1.25 needs to be added for low mains operation, commutation delay, internal drops etc. We shall come back to this later. Hence $U_{line} = 280 \text{ V}$.

Therefore, maximum ripple input = $280 \times \sqrt{2} = 395$ volts at 300 Hz pk-pk.

The magnet will give an attenuation of $\frac{300}{0.16} = 1875$ at 300 Hz

$\frac{50}{0.16} = 313$ at 50 Hz.

Hence ripple attenuation of magnet at 300 Hz = $\frac{395}{1875} \times \frac{1}{300} = 7 \times 10^{-4}$.

At 300 Hz we need an additional factor of $\frac{5 \times 10^{-4}}{7 \times 10^{-4}} = 7.1 \times 10^{-1}$.

If we take an LC circuit then

$\left(\frac{f_0}{f_{300}}\right)^2 = 7.1 \times 10^{-3}$

$f_0 = f_{300}\sqrt{7.1 \times 10^{-3}}$

$= 300 \times 0.084$

$= 25$ Hz.

The vacuum chamber attenuation will give a factor of 10 (hence flux ripple = $5 \times 10^{-7}$). Further, the fundamental of the bridge output is less than 395 V pk-pk.

We don't need an active filter for ripple rejection. But do we need one for mains changes? Take the critical 1 Hz region:

- magnet attenuation $= \frac{1}{0.16} = 6.25$;
- converter rejection $= 120$;
- passive filter rejection $= 1$.

Hence a 10% mains change becomes: $\frac{0.1}{6.25 \times 120} = 1.3 \times 10^{-4}$

We are just about there. Although the inclusion of an active filter at this stage may not be evident our design might like to take into consideration its later addition. So $f_0 = 1/2\pi\sqrt{LC}$,

$LC = \frac{1}{(2\pi f_0)^2} = \frac{1}{(2\pi \cdot 25)^2} = 4.4 \times 10^{-5}$

At first sight we can choose any value of $L$ and $C$ satisfying the above. However, very large 'C' is impossible to obtain because of the ESR while low 'L' would give high peak currents. The contrary would also be true.

To get an answer we need to look at the economies. Since we have an undamped LC filter, if all electronic damping fails, we could have above 600 V from an accidental step input. Hence, if we want to avoid series connection of electrolytics, we must consider paper, metallized paper or polypropylene capacitors.
Now the total cost of the filter is \( P_L + P_C = P_T \). If \( P_L = aL \) and \( P_C = bC \) and \( LC = d \), then
\[
P_T = aL + bC = \frac{ad}{C} + bC
\]

Minimum cost when \( \frac{\Delta P_T}{\Delta C} = 0 = -\frac{ad}{C^2} + b \)

\[
\therefore \quad C = \sqrt{\frac{ad}{b}} \quad \text{for minimum cost}
\]
\[
L = \sqrt{\frac{bd}{a}}
\]
\[
P_C = b \sqrt{\frac{ad}{b}} = \sqrt{ab} \quad \text{d}
\]
\[
P_L = a \sqrt{\frac{bd}{a}} = \sqrt{ab} \quad \text{d}
\]

\[
\therefore \quad P_L = P_C
\]

As would be expected, minimum cost occurs when the cost of the choke equals the cost of the capacitor:
- cost of paper capacitors \((P_C) \sim 287 \text{ (U)} 1.15 \cdot C \cdot \text{SF}\);
- cost of chokes \((P_L) \sim 3.5 \text{ L (SF)}\).

Most paper or polypropylene capacitors can take an over voltage of about 38% for a small time.

Worst case peak D.C. voltage = \( 1.35 \times U_{\text{line}} \times 2 = 756 \text{ V} \).

Taking into consideration the over-voltage rating above a capacitor of about 600 V nominal would be sufficient. Hence:

\[
P_C = 287 \cdot (600)^{1.15} \cdot C = 4.5 \times 10^5 \cdot C
\]

and

\[
P_L = 3.5 \cdot (300)^2 \cdot L = 3.15 \times 10^5 \cdot L
\]

\[
\therefore \quad P_L = P_C
\]

\[
\therefore \quad C = \frac{3.15 \times 10^5 \cdot L}{4.5 \times 10^5} \quad \text{and} \quad L = \frac{4.40 \times 10^{-3}}{C}
\]
\[
C = \sqrt{0.7 \times 4.4 \times 10^{-3} \, \text{mF}} \approx 5.5 \text{ mF} \quad \text{at 600 V}
\]

and
\[
L = \frac{4.4 \times 10^{-3}}{5.5 \times 10^{-3}} = 8 \text{ mH at 300 A}
\]

If we require natural damping of the filter, then this is normally done by splitting the capacitor in two and adding a damping resistor to the large capacitor (Fig. 9). In this case a satisfactory response is found when

\[
C_1 = 5C_2
\]

In our case

\[
C_1 = 4.6 \text{ mF}
\]

and

\[
C_2 = 0.9 \text{ mF}.
\]

The exact size will depend on the market availability. If we want to limit overshoot to 1.5 times nominal, then

\[
R = 0.4 \sqrt{\frac{L}{C_1}} = 0.4 \sqrt{\frac{8}{4.6}} = 0.53 \Omega
\]
Attenuation at 300 Hz would now be about 12 dB less (factor of 4). This would either absorb some of our vacuum chamber safety factor or we would need a larger filter. However, considering the components we have then the ripple voltage at the output terminals is given by:

\[
\text{Input ripple} \times \text{attenuation of filter (input ripple = 395 \times 0.71)}
\]
\[
= 280 \times \frac{1}{40} \text{ pk–pk}
\]
\[
= \frac{1}{2} \times \frac{1}{\sqrt{2}} \times 280 \times \frac{1}{40} \text{ r.m.s.}
\]
\[
= 2.47 \text{ volts r.m.s.}
\]
\[
\therefore \text{current in branch } R \ C_1 = \frac{2.47}{Z} = \frac{2.47}{\sqrt{R^2 + X_C^2}} = \frac{2.47}{0.54} = 4.6 \text{ A}
\]

Hence, continuous dissipation = \((4.6)^2 \times 0.53 = 11.2\) watts. This is a very small power. In reality, the resistor must withstand the shock of a pulse charge (0 to 450 V) without damage, as well as an accidental oscillation until the detection circuits work. This means a peak-peak oscillation of 1.5 times nominal for about 10 seconds.

\[
\text{Peak power} = \left( \frac{\text{Ur.m.s.}}{Z_m} \right)^2 \cdot R
\]
\[
Z_m = \sqrt{0.53^2 + 1.44^2} = 1.53 \text{ } \Omega
\]
\[
\text{Ur.m.s.} = \frac{300 \times 1.5}{2 \times \sqrt{2}} = 159 \text{ volts r.m.s.}
\]
\[
\therefore \text{Peak power} = \left( \frac{159}{1.53} \right)^2 \times 0.53 = 5.7 \text{ kW for 10 seconds.}
\]

This is a much more severe specification. Likewise \( C_1 \) must withstand 4.6 A r.m.s. continuously and 104 A r.m.s. for 10 seconds at 25 Hz.

For \( C_2 \) we have:

\[
I_{c_2} = \frac{\text{Uripple}}{X_{C_2}} = \frac{2.47}{0.59} = 4.19 \text{ A r.m.s.}
\]

continuously and worst case current:

\[
= \frac{159}{X_{C_2}} = \frac{159}{7.37} = 21.6 \text{ A r.m.s. for 10 seconds}
\]

Now try a simulation to see if the figures are correct!

6.4 Thyristor Bridge

The average current through each thyristor is simply one third of the D.C. since it conducts for 120° as an approximate square wave supplying an inductive load. The power dissipated can be taken from the manufacturer's data sheets, but if we assume about 1.3 V forward voltage drop we have:

\[
I_{T(\text{AV})} = \frac{300}{3} = 100 \text{ A}
\]
\[
\text{Power} = 100 \times 1.3 = 130 \text{ watts.}
\]

To operate with 40°C ambient and a junction temperature which does not exceed 125°C, we need to achieve
a total temperature coefficient of less than:

\[
\frac{85}{130} = 0.65^\circ C/watt.
\]

Since the device will have a junction to case coefficient of 0.12\(^\circ\)C/W, then we need a heat sink of 0.53 kW. A naturally air-cooled sink of about 1.2 litres volume will do this.

Since we prefer to leave away fuses, we normally design so that the thyristors can withstand a large surge. The r.m.s. current in the transformer secondary is

\[
\frac{\sqrt{2}}{\sqrt{3}} \times 300 = 245 \text{ A}
\]

A typical transformer would be around 5 - 6% impedance. Hence its peak 1/2 wave current for a symmetric fault would be:

\[
20 \times 245 \times \sqrt{2} = 7 \text{ kA}.
\]

Although not representing the worst-case fault current that the thyristor can be subjected to, this figure does represent a compromise giving a reasonable rating to the thyristor without being too large. A thyristor having an \(I_{TSM}\) of 7 kA for 50% re-applied voltage would typically have an average current rating in the region of 200 to 250 A. Hence, it would be very conservatively rated for its continuous operation but we do away with the fuses which are expensive and can be unreliable.

Our designs are based on the idea that a thyristor should never need to be replaced and hence not only is the current rating conservative but also the voltage rating. We normally apply a factor of 3.7. Hence,

\[
V_{RRM} = 280 \times 3.7 = 1040 \text{ V}, \text{ say 1200 volts grade thyristor.}
\]

It is also vital to supply good but simple voltage suppression [8] around the thyristors, not only for the thyristors but other sensitive devices.

We use two types, each doing a different job:
1) RC networks across each thyristor to take care of the reverse recovery during commutation;
2) bucket reservoirs to take care of the magnetic energy in the transformer.

A simple capacitor is placed across each thyristor to suppress the reverse recovery voltage to well within the voltage rating of the device. We need to know the rate of commutation which depends on the transformer reactance.

This can be calculated from the transformer impedance. In our case:

\[
2\pi f L = \frac{U_{\text{phase}} \times 5}{I_{\text{phase}} \times 100}
\]

\[
\therefore L = \frac{V}{1 \times 20 \times 2\pi 50} = \frac{280}{245 \times 10^3 \times 2\pi} = 181 \mu\text{H}
\]

The commutation inductance is \(2 \times 181 \mu\text{H} = 362 \mu\text{H}\) for two branches:
\[ \frac{\Delta I}{\Delta T} = \frac{U}{L} = \frac{395 \text{ V}}{362 \mu\text{H}} = 1.1 \text{ A/\mu s} \]

Now the data sheets need to tell us the stored charge at the worst case of 125°C. This might be 700 µC at 1.1 A/µs for the chosen thyristor.

For an overshoot factor of 1.7 we have:

\[ C = \frac{0.4 \times Q}{\text{Ur.m.s.}} = \frac{0.4 \times 700 \times 10^{-6}}{280} = 1 \mu\text{F} \]

Voltage rating should be similar to the thyristor, i.e. 1000 V. A resistance should be installed to damp oscillations

\[ R = \sqrt{\frac{L}{C}} = \sqrt{\frac{362}{1}} = 19 \Omega \text{ say 22 } \Omega \]

Power = \[200 \times C \times U^2_{\text{r.m.s.}} = 200 \times 10^{-6} \times (280)^2 = 15 \text{ watts.} \]

The bucket network consists of a full-wave diode bridge feeding a reservoir capacitor (Fig. 8). The capacitor should absorb the energy in the transformer. This is 1/2 L[t]:

\[ = \frac{1}{2} \times 362 \times 10^{-6} \times (300)^2 \]

\[ = 16 \text{ joules.} \]

Doubling this figure to account for an upstream supply transformer gives 32 joules.

We already have \[280 \times \sqrt{2} \times 1.1 \text{ volts on the capacitor which gives 435 V.} \] We would like to limit the maximum voltage to say 1000 V. Hence:

\[ 32 = \frac{1}{2} C \left(1000^2 - 435^2\right) \]

\[ \therefore C = 78 \mu\text{F} \]

Hence, a 78 µF, 1200 V capacitor will do. The discharge resistor should have a time constant of about 100 ms. Therefore \[R_d = 2.2 \text{ k} \Omega.\] The series resistance should damp the circuit and limit the initial charge. Using 1/2 critical damping gives:

\[ R_s = \sqrt{\frac{L}{C}} = \sqrt{\frac{362}{78}} = 2.2 \Omega \]

This gives a peak current of 200 A at switch on.

6.5 The Transformer

By default we already have many of the parameters:

- \[U_{\text{secondary}} \text{ line-line} = 280 \text{ volts at full load} \]
- Impedance voltage = 5 - 7%
- \[I_{\text{secondary}} = 245 \text{ A r.m.s. square wave} \]
- \[U_{\text{primary}} \text{ line-line} = 380 \text{ V r.m.s. 50 Hz.} \]
There are however a few additional points. The secondary current is a square wave and although 245 A is the mathematical r.m.s. value, the windings will suffer additional losses due to the higher frequency components. An equivalent pure sine-wave of 1.1 to 1.2 larger may have to be used to get the equivalent kVA and also for test purposes. Hence we need a 130 kVA transformer to achieve 90 kW output. The power factor as seen from the mains will be 0.76 however.

In order not to disturb the network at switch on and also not to cause trips, the in-rush current should be limited or a pre-charge circuit used. In this case, the peak current of the first half cycle should not exceed $10 \times I_n$:

$$
I_n = \frac{90000}{0.76 \times 380 \times \sqrt{3}} = 180 \text{ A r.m.s.}
$$

$I_{\text{inrush}} < 1800 \text{ A peak}$.

The configuration should have at least one Delta winding. Temperature class would be 'B' or better with a hotspot not to exceed 120°C with 40°C ambient. Temperature rise by resistance should be limited to 55 K with natural air cooling.

Since this design does not use fuses it is important that the transformer is constructed and type tested for both symmetric and asymmetric short-circuit.

6.6 The Circuit-Breaker/Input-Switch

The network impedance is the most important factor for this element. Assume a 2 MVA transformer is used having 6% voltage impedance. This gives 50 kA short-circuit current. The device would thus be specified:

- short-circuit breaking capacity : 50 kA r.m.s. 50 Hz (cat. P1)
- short-circuit breaking capacity : 20 kA r.m.s. 50 Hz (cat. P2)
- magnetic overload range : 1600 A - 2500 A
- thermal overload range = 160 A - 250 A
- 10000 operating cycles with 1800 A inrush current.

The P1 category requires that the device be replaced after such a fault while P2 allows it to be used again. In reality, the supply cables would limit the short-circuit current to much nearer the P2 value.

7. CONCLUSION

Achieving high performance in power converters requires several compromises to be made if an overall acceptable performance is to be attained. Make sure the requirements are correctly and fully elaborated with the users before starting a design. Although the more glamorous aspects are more interesting to work on, do not forget the all important reliability and availability needed in modern accelerator operation. Take the simplest well engineered design to meet your specification and remember that the power converters are a relatively inexpensive part of an accelerator while being vital to its well being. Avoid false economies that could destroy this.
ACKNOWLEDGEMENTS

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POWER CONVERTER FEEDBACK CONTROL

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Abstract
The power properties and design criteria of a power converter feedback system are described. From the electrical model, operational characteristics are deduced. Based on the above, as well as the specification and measurement problems, the control system is defined. This leads us to a cascaded voltage loop and current loop, on which control synthesis is applied using various approaches generated by the different specification requirements. A novel regulation principle for a series-connected power converter is also described.

1. INTRODUCTION

In the domain of accelerators, power converter performance has progressed notably during the last two decades. Whether it be conventional equipment or the more modern generation of power converters the fundamental objectives, other than electrical problems, have always been the precision and the reliability.

The elements concerned most directly with these objectives and more importantly towards the precision are:
- the reference source;
- the current transducer;
- the regulation circuits.

The references of the potentiometer type driven by up-down motors have been replaced by digital to analogue converters, which have greatly improved over the last years and can now fulfil most of our applications.

Likewise the current transducers have followed a similar trend and, along with the galvanic isolation of the DCCT, once again fulfil our requirements.

Hence, a particular attention should now be paid to the regulation loops which were previously masked by the lower performance of the former elements. Further, the elements such as operational amplifiers etc, used as the fundamental building blocks of the regulation system no longer form the limiting factor of performance.

The domain of regulation as well as the numerous methods of analysis being vast [1, 2], we have been obliged to limit the content of this course. Our objective therefore is to follow a rapid overview of the various elements of a power converter which will lead us to a regulation topology having all the qualities required of a reliable high precision converter for use in particle accelerators.
2. **DESCRIPTION OF A POWER CONVERTER AND ITS LOAD**

2.1 **The Electrical Model**

We can sub-divide the power converter into four principle blocks:

![Block diagram of a power converter](image)

**Fig. 1** Block diagram of a power converter

- **Adaptation** of the network to the user needs is normally carried out using a transformer which also provides galvanic isolation.
- The **regulation** and **rectification** can be carried out by a phase controlled thyristor bridge driven by a suitable firing circuit.
- The **filter** must reduce the residual ripple to an acceptable level for the user. It is normally of the type shown in figure 2 or a simple LC filter.
- The **load** consists of a connection between the converter and its load as well as the resistance and inductance of the latter (Fig. 3). Other parasitic effects such as capacitance to earth may need to be taken into consideration. While the load is always the last element, the other series-connected elements can be in any order, and of any type (e.g. thyristor control or switch mode).

![Filter module](image)

**Fig. 2** Filter module

![Equivalent diagram of a magnet load](image)

**Fig. 3** Equivalent diagram of a magnet load

2.2 **Operation Characteristics**

Before going onto the regulation, we should first consider the characteristics of the three fundamental elements concerned in the command process, i.e. the thyristor bridge including its transformer, the passive filter and the load [3].
2.2.1 The Thyristor Bridge

Consider figure 4 which shows the output of a full-wave thyristor bridge. In the case of a balanced network, the average DC output as a function of the firing delay $\alpha$ is given by:

$$\bar{V}_p = \frac{3}{\pi} V_o \cos \alpha$$

(1)

![Diagram](image)

Fig. 4 Bridge output voltage

This characteristic is shown in figure 5 which shows the case of discontinuous conduction. The transfer function of $\alpha$ to $\bar{V}_p$ is therefore non-linear.

![Diagram](image)

Fig. 5 Static characteristic of the bridge
In order to avoid the problems associated with this non-linear characteristic we normally use a firing circuit which performs linearization of this characteristic. For instance the use of a closed-loop firing circuit associated with the thyristor bridge can give a practically linear static relationship as shown in figure 6 between the input reference voltage and mean bridge output voltage. Instability may result below 1% which can be solved by the addition of a current bleed, which amounts to an offset in the characteristic.

The dynamic response is shown in the Bode plot of figure 6 where it can be seen that above 75 Hz the response starts to roll-off rapidly. This process cannot be compensated for in a control loop, and gives the fundamental frequency limitation of the system.

a) **BLOCK DIAGRAM OF A CLOSED-LOOP FIRING CIRCUIT AND A THYRISTOR BRIDGE**

![Block Diagram](image)

b) **STATIC CHARACTERISTIC**

![Static Characteristic](image)

c) **BODE DIAGRAM**

![Bode Diagram](image)

Fig. 6 Closed-loop firing circuit of a thyristor bridge
2.2.2 Passive Filter

The parameters, resonant frequency and damping, of the passive filter are dictated by the residual ripple tolerated by the load. We have two possibilities:

1) place the resonance of the filter above or close to the frequency limit of the command element (e.g. 75 Hz for a thyristor bridge). The consequences are that no compensation can be obtained for the filter and in particular correct damping cannot be achieved electronically. Therefore, the filter may need to be damped electrically by adding a damping resistor and eventually a second C and the attenuation may not be sufficient;

2) place the resonance of the filter sufficiently below the limit of the command element such that compensation can be achieved. Hence, the filter can be undamped which will give higher attenuation with no extra cost. The large damping resistor is no longer needed leading to further economies.

Once the introduction of a passive filter is necessary, it is evident that the ratio attenuation/price becomes very attractive if we can choose a filter below the bandwidth limit of the system. To illustrate this we will consider the case of an LC filter at 90 Hz and 30 Hz used on the output of a 6-pulse thyristor bridge having a band pass limit of 75 Hz.

The load impedance can be ignored compared with the output impedance of the filter giving a transfer function:

\[
\frac{V_c(s)}{V_p(s)} = \frac{1 + RCs}{1 + RCs + LCs^2}
\]  

For a filter at 90 Hz, we have:

\[\frac{L}{C} = \frac{1}{(2\pi \times 90)^2}\]  
e.g. \(L = 1 \text{ mH}, C = 3 \text{ mF}\)  

(3)

A reasonable value of damping is given by:

\[R = \frac{\sqrt{L}}{\sqrt{C}}\]  
e.g. \(R = 0.58 \Omega\)

(4)

which is half the mathematical critical damping.

For the filter at 30 Hz, we have:

\[\frac{L}{C} = \frac{1}{(2\pi \times 30)^2}\]  
e.g. \(L = 3 \text{ mH}, C = 9 \text{ mF}\)

(5)

We no longer need a damping resistor but must make sure that any parasitic resistance (e.g. connections, capacitor ESR) does not hinder the correct operation of the filter, i.e. gives a zero in equation (2) above 300 Hz.

\[R \leq \frac{1}{2\pi \times 300 \times C}\]  
e.g. \(R \leq 59 \text{ m\Omega}\)

(6)
Looking at the Bode plots of figure 7, we can see a difference of attenuation of 35 dB (factor of 56) at 300 Hz; this improvement for a cost increase of about 3.

Fig. 7 Bode plot of filters

2.2.3 Load

The load is the most important element of the system since it dominates the specification. Its make up is simple since it is normally a first-order element \((0.1 < \tau < 100 \text{ s})\). However, a non-linear characteristic can result due to saturation of the magnetic circuit. In certain cases coupling with the vacuum chamber or a shorted-turn caused by a cooling pipe can require a higher-order model.

3. STRUCTURE OF THE CLOSED-LOOP SYSTEM

Having defined the fundamental elements of the system, we can now consider the structure of a closed-loop control system. However, we must also take into consideration the quality of the various feed-back signals we need and the specification of the power converter itself [4, 5].

3.1 Specification

The following are the points of particular interest from a point of view of the closed-loop control:

- precision;
- resolution;
- long-term drift;
- perturbation error;
- following error;
- overshoot.
The precision defines generally the absolute tolerance of the current. It relies on the quality of the reference and the current transducer which must not be degraded by the control loops.

The resolution depends essentially on the reference, since the rest of the process has no element likely to influence the resolution.

The long-term drift depends essentially on temperature and ageing effects on the reference and current transducer and to a lesser extent those of the regulation electronics.

The perturbation error is the most critical point. Firstly there is the continuous ripple resulting from chopping the 3-phase mains current to form the bridge output. In general the passive filter has sufficient rejection for this type of perturbation. If this is not the case, we may need to add an active filter. Secondly, there is the random mains perturbation which has low-frequency components for which the passive filter has little rejection. This is often the most usual reason to add an active filter.

Fig. 8 Illustration of the precision long-term drift and perturbation error

The following error is only applicable to a dynamic change and depends entirely on the quality of the loops.

The overshoot specification comes from the hysteresis limits of the magnet as well as the obvious beam requirements in this sense.

3.2 Current measurement

Current measurement influences directly the precision, long-term drift and to a lesser extent the perturbation rejection. The performance of the system depends to a large extent on this element. The most usual device today is a zero-flux type of DCCT (Direct Current Current Transducer) although the older Hingorani type can still be found. The limiting performance for us is the induced noise into the primary bus-bar and noise on the output signal. These two effects are illustrated in figure 9.
The noise on the signal is particularly annoying since it will appear throughout the control process within the bandwidth of the system.

The induced noise in the bus-bars can only be eliminated by including it within a loop of sufficient bandwidth such as an active filter.

The zero-flux type offers better performance in both these domains.

The electronics employed in the zero-flux type now give a comparable reliability to the older Hингорani type while the basic concept gives a far superior performance.

3.3 Formation of the control signal

The block diagram (Fig. 10) represents the principle elements of the control process and the perturbations likely to create a fluctuation of the output current.

![Diagram](image)

**Fig. 10** Block diagram of control process and perturbations

The perturbations come from three sources:
- the mains network;
- the current transducer;
- the load.

The typical effects of a mains perturbation and that of the current transducer on the output voltage are shown in figure 11, photos a and b.
Photo a - Transient perturbation of 2.5%  

Photo b - Perturbation due to Hingorani current transducer

Fig. 11  Perturbation effects on the output voltage

The perturbations introduced by the load are mainly due to resistance variations but can also be due to coupling with other sources.

The rejection of the mains perturbations as seen on the output voltage can be achieved in two ways:

1) measure the current and correct at the input;
2) measure the voltage on the output and correct at the input.

In order to resolve this problem, we would look at figure 12, where the typical analysis of a filter as well as the magnet is given.

Fig. 12  Bode plot of filter and magnet
It is evident that the information contained in the voltage is much richer than that of the current for bandwidth above the magnet break-point. Hence, the second method consisting of generating a correction signal from the output voltage is much more interesting particularly when we take into consideration the signal-to-noise ratio.

In simpler terms it appears evident that one should correct the source of the perturbation as soon as it can be measured rather than correct a secondary effect. In particular, a simple voltage divider can be used to provide the required information.

However, the current being the main control parameter, it is from the current transducer that we shall establish the low frequency command of the system as well as the rejection of the load perturbations.

Together this leads to the principle diagram shown in figure 13 where we control the output voltage (voltage loop) and the output current (current loop).

![Principle block diagram of the control system](image)

**Fig. 13 Principle block diagram of the control system**

The compensation of the voltage loop generates a command signal for the firing system thus producing a near perfect voltage source at the output terminals.

Likewise the compensation of the current loop associated with the load inductance forms a current source supplied by the previous voltage source. Thus the rules of cascaded voltage and current sources are rigorously respected.

4. **CONTROL SYNTHESIS**

4.1 **Definition of the performance criteria**

In general, for the type of control that interests us as shown in figure 14, we wish to minimise error between the reference \( u \) and the output \( y \) by adjusting the regulation parameters of the compensation.
Fig. 14 Block diagram of a closed-loop system

The behaviour of the system in relation to perturbations is governed principally by its dynamic response. It is therefore important to assure that this is adequate by a choice of compensation. Since the dynamic behaviour can be easily observed by the step response of the system, we have chosen this method to compare the relative responses of different compensations. In this respect we have chosen to minimize the integral squared error (ISE) defined as:

\[
ISE = \int_0^T e^2(t) \, dt
\]

where \( e(t) \) represents the error as shown in the shaded area of figure 15, with a step input introduced at \( t = 0 \) and a perturbation at \( t = t_p \).

Fig. 15 Error evolution

In an ideal case ISE = 0, but in reality ISE is always greater than zero. We need a compensation which will minimize ISE. For a second order system this leads us to maximize the pulsation, \( \omega_0 \) and to a damping factor \( \zeta \) of 0.5 [2].

4.2 Voltage loop

The objectives of the voltage loop are to provide satisfactory mains rejection while preserving the dynamic performance criteria we have chosen.

Since the load is decoupled, we can simply consider the thyristor bridge and passive filter in an autonomous way. If the passive filter has a minimum damping factor \( \zeta \) of 0.2, then a proportional integrator (PI) compensator associated with a phase advance will give the required performance as seen in figure 16. There we can see that the bandwidth has practically doubled while the damping factor has increased to the desired level.
Photo a - Step response with voltage-loop  

Photo b - Step response without voltage-loop

Fig. 16 Comparison of performance with and without voltage loop

However, if we wish to achieve the same performance from an undamped filter, which has been chosen for its superior attenuation, then this type of compensation becomes insufficient. We would then turn towards state feedback compensation, the principles of which will be recalled.

4.2.1 State feedback

Consider a linear system represented by a state model in the form of:

\[
\frac{dX}{dt} = A \cdot X(t) + B \cdot U(t) \\
Y(t) = C \cdot X(t) + D \cdot U(t)
\]  \hspace{1cm} (8)

where

- X(t) is the state vector of dimension n,
- U(t) is the input vector of dimension p,
- Y(t) is the output vector of dimension q,

and (A, B, C, D) are the state matrices of the system.

The matrix A is the dynamic matrix of the system. Its eigen values characterize the dynamics of the system. The principle of the state feedback uses a law of the form:

\[
U(t) = E(t) + K \cdot X(t)
\]  \hspace{1cm} (9)

The matrix K is called the feedback matrix and has a dimension of p x n.

The state equation of the compensated system is:

\[
\frac{dX}{dt} = (A + B \cdot K) \cdot X(t) + B \cdot E(t) \hspace{1cm} (10)
\]

\[
X(t) = (C + D \cdot K) \cdot X(t) + D \cdot E(t) \hspace{1cm} (11)
\]
The new dynamic matrix is thus $A + B \cdot K$. It can be seen that the correct choice of the matrix $K$ will improve the stability and the dynamics of the system.

One can show that the idea of governability of a system is equivalent to the ability to arbitrarily choose the mode of the closed system [6 - 8].

The matrix $K$ can be determined in a systematic way (hence simply programmed) from a knowledge of $A$, $B$ and $(A + B \cdot K)$.

There exists a number of cases where all the states of the system are not available or measurable. However, there exists certain methods of deriving these from the input and outputs of the system (observers and dynamic compensators) [2, 9, 10].

4.2.2 Application of state feedback to the passive filter

In order to illustrate the use of state feedback, we will consider the passive filter of figure 17.

Fig. 17 Passive filter

It can be represented by:

$$
\begin{bmatrix}
\frac{dI}{dt} \\
\frac{dV_c}{dt}
\end{bmatrix} =
\begin{bmatrix}
0 & -1/L \\
1/C & -R/L
\end{bmatrix}
\begin{bmatrix}
I \\
V_c
\end{bmatrix} +
\begin{bmatrix}
1/L \\
R/L
\end{bmatrix} V_r
$$

(12)

We then define a state feedback applying a law of the form:

$$
V_r = U + K X = U + [k_1, k_2] [I \ V_c]^T
$$

(13)

The dynamic matrix of the system becomes:

$$
A + B \cdot K =
\begin{bmatrix}
k_1/L & (k_2 - 1)/L \\
1/C + k_1 R & (k_2 - 1) R/L
\end{bmatrix}
$$

(14)
Thus we deduce the characteristic equation:

\[
\frac{L}{1-k_2} C s^2 + \left( R - \frac{k_1 C}{1-k_2} \right) s + 1 = 0
\]  

(15)

from which we can see the possibility of adjusting the resonant frequency and the damping factor. For instance if we would like to double the bandwidth, we would have \(k_2 = -3\), and we can choose the damping factor \(\zeta\) this giving:

\[
k_1 = 4 \left( R - \zeta \sqrt{\frac{L}{C}} \right)
\]  

(16)

For our example, we can take the capacitor current and the output voltage as state variables neither of which are difficult to measure. Although this type of correction can determine the dynamic properties of the system, it does not eliminate the static error. For this we must add an integral action on the output voltage. Reality has shown that it is more interesting to introduce a proportional action instead of the state feedback \(k_2\).

While the above reasoning has not taken into account the thyristor bridge, this is of little importance for small signal operation. However, our system must be able to respond to large amplitude signals from the reference. In this case, we fall on the limit of linear operation of the thyristor bridge, and the voltage source is temporarily lost. In order to remedy this situation we can introduce a slowing down of the system. Since we do not wish to deteriorate the dynamics of the system, we have chosen to introduce a small phase advance compensator in the feedback which gives a filtering effect seen from the reference. The system is therefore shown in figure 18 and the performance thus obtained in figure 19.

Fig. 18 Block diagram of the voltage loop
4.2.3 Introduction of an active filter

There are limits to the performance that can be achieved with a passive filter alone and once these are reached we must introduce an active element [11]. The choice of the active element as well as its command structure is a vast subject which we will not cover in detail. We will however consider the general principles, developed around the system shown in figure 20.

The system now has two inputs which demands a different strategy to that used previously. The reasons for introducing the active element are not only to reduce the residual ripple but perhaps more importantly to increase the performance of the system at the lower frequencies where the passive filter has little or no effect. This leads to the structure shown in figure 21. The vital element is the feedback matrix $K$ which, from the state vector of the filter, generates the command for the two inputs in a manner to render the two compatible and give a resultant single input system with increased bandwidth. The rejection curves shown in figure 22 demonstrate the successive improvements made when adding an active filter. The curve (a) has no active filter. The curve (b) shows the addition of the active filter purely as a filtering element. The curves (c) and (c') show the rejection when the active filter is used also to increase the low-frequency gain of the system. An improvement of 50 dB at 10 Hz and 20 dB at 1 Hz can be seen.
Fig. 20 Active element in the capacitor branch

Fig. 21 Voltage loop with active filter

Fig. 22 Rejection curves
4.3 Current loop

We have thus obtained, one way or the other, a suitable voltage source as input to our current loop. For the current loop we need to consider two distinct modes of operation:
1) utilization of the converter in a static state (Beam transfer, flat-top operation);
2) the converter used in an accelerator where, as well as the static requirements, it must ramp during acceleration.

In the two cases the precision obtained is the main objective.

4.3.1 The static case

In this case a zero static error is needed which can be obtained by a PI compensator. The system could be represented by the block diagram of figure 23.

![Block diagram of current-loop](image)

Fig. 23 Block diagram of current-loop

The current transducer is included with the load as well as its internally generated noise.

The main points to take into consideration in this case are:
1) the bandwidth required in the current loop to avoid:
   a) interference with the voltage loop;
   b) noise injection from the current transducer;
   c) noise from reference source, particularly if it is not close to the control loop;
2) how to reject the resistance changes of the load and cope with the inductance changes;
3) what type of regulation should we adopt to minimize the overshoot to a step input?

Concerning the interference with the voltage loop, we should maintain a margin of about one decade below the closed voltage loop.

The influence of the current transducer noise depends on the gain of the system at the noise frequency. The use of a zero-flux transducer has shown that in most cases it represents no limitation for the system. Concerning the reference source noise, it is clear that the slower the system the less will be its influence. Taking into consideration these points, practical experience has shown that a bandwidth of a few Hertz represents an interesting compromise.

Concerning the rejection of resistance change of the load, it should be noted that during a change of temperature, following a step-change in current, the simple integrator cannot reject this completely, although the effect in practice is very small, since the rate of change of resistance is small for most magnets.
The problems of overshoot can be resolved principally in two ways:

1) introduction of a zero in the feedback of the compensation.
This is equivalent to placing a filter on the reference. However, this must be associated with a pole in order to avoid noise at the summing point. The resulting process is shown in figure 24.

![Diagram](image)

Fig. 24 Introduction of a zero in the feedback

If the bandwidth of the closed-loop system is large compared to the position of the pole, it is evident that the main mode will be the one of the pole and therefore the overshoot problem will be solved.

This method has the inconvenience of hiding the closed-loop systems such that any instability during a step test may not be seen.

2) The second solution is inspired by the first but consists of a filter introduced in the reference which can be easily removed for dynamic tests.

This method, associated with a limiter to avoid saturation during large reference changes, leads to the block diagram shown in figure 25.

![Diagram](image)

Fig. 25 Block diagram of current loop with limiter

The limiter is a non-linear element with two thresholds responding only to large reference changes and acting on that reference. It also gives minimum and maximum operating levels of the voltage source.
The performance obtained with this type of system is shown in figure 26.

![Graphs showing performance](image)

**Photo a** - Response within the linear domain without limiter and filter
**Photo b** - Response to large step with limiter and filter

Fig. 26 Performance evaluation

4.3.2 **The ramping case**

Concerning the utilization of a power converter in an accelerator, it may be necessary to modify the static constraints as a result of the problems involved in following a dynamic reference [12]. For the current loop these are:
- increase of the bandwidth to a maximum possible in order to reduce transient times;
- introduction of a double integration to give zero steady state following error.

These two points will introduce new stability problems as well as different transient behaviour. Of course the previously added reference filter needs to be removed unless we can assure its high precision.

The increase of bandwith can only be achieved to the detriment of noise, which puts more severe constraints on the current transducer, reference source and other noises. The double integration is necessary since a simple integration gives a following error of:

\[ e_f = \frac{1}{K} \]  \hspace{1cm} (17)

where K is the gain of the open loop which is a function of the compensation components and the resistance of the load. Any change of components will therefore modify the following error and resistance changes will also produce a non-negligible variation of this error.

In choosing between single or double integration the time during which steady-state ramping takes place may well dictate which is to be used. If this time is short compared with the transient periods, then the added complication of double integration may not be justified.
4.3.3 **Overshoot**

In the case of a dynamic reference, the problems of overshoot can be resolved by the methods used before. Under these conditions we introduce a following error which we can compensate by modifying the time at which the ramp is started. However, the changes in the following error caused by the changes in the components of the filter cannot be taken into account in a systematic way.

The ideal method that one could suggest to eliminate this type of problem is to act numerically on the reference source in order to generate the appropriate signal to eliminate all these errors, calculated from the knowledge of the system and the desired output function.

5. **MASTER-SLAVE SYSTEM**

We will consider in this part the control of several power converters placed in series, via their loads, around the circumference of a machine [13]. Such a connection is frequently found in the chains of the principle dipoles and quadrupoles in order to distribute the power and limit the voltage to earth.

It is evident from the most elementary rules that the regulation of each unit in current is prohibited. We are therefore obliged to consider the type of structure shown in figure 27.

![Block diagram of master-slave with liaison element](image)

**Fig. 27** Block diagram of master-slave with liaison element

The current compensator drives in parallel the slave voltage sources, which we consider to be ideal, via a liaison element (cable, communication system, etc). If the latter is ideal, then the resulting system seen by the current compensator is the same as a single system.

In fact, only the gain is modified by the number of power converters.

Once the distances between the master and slave converters becomes large (e.g. 14 km in LEP), then the choice of liaison element becomes critical from both a technical and economic point of view. The solution of digital transmission of references to the slave elements generally results in a non-linear system. We can eliminate the liaison element by taking the necessary information where it exists naturally. That is to say the current in the
series-connected magnet. This leads us to a block diagram as shown in figure 28 from which we will derive the necessary compensation needed for each slave power converter.

![Block diagram of master-slave without liaison element](image)

**Fig. 28** Block diagram of master-slave without liaison element

**Determination of the required compensation**

The system composed of 'n' power converters in a configuration of master-slave without liaison element is represented by the functional diagram of figure 29.

![Function diagram of 'n' power converters](image)

**Fig. 29** Function diagram of 'n' power converters

\[
\begin{align*}
H_s & : \text{ transfer function of the load} \\
 k_i \{1 \ldots n\} & : \text{ current transducer gain} \\
 H_i \{1, \ldots n-1\} & : \text{ transfer function of slave voltage sources} \\
 H_n & : \text{ transfer function of the voltage source of the master} \\
 H_c & : \text{ transfer function of the current loop compensator.}
\end{align*}
\]

The problem we have to resolve is whether we can compensate the slave converter in such a way that the input to the master converter shares the static and dynamic characteristics between all converters.
In other words, if we have for the load:

\[ H_L \equiv \frac{1}{R(1 + L/R \cdot s)} \]  

(18)

We should have seen from the input of the master converter:

\[ \frac{I(s)}{V_C(s)} = \frac{H_L}{R \left( 1 + \frac{L}{R \cdot n} s \right)} \]  

(19)

That is to say that the resistance and the time-constant must be divided by \( n \) in order to meet our criteria. From the functional diagram we obtain:

\[ \frac{I(s)}{V_C(s)} = \frac{H_L(s) \cdot H_i(s)}{1 - H_L(s) \sum_{i=1}^{n-1} H_i(s) \cdot k_i} \]  

(20)

We must determine \( \sum_{i=1}^{n-1} H_i(s) \cdot k_i \) such that:

\[ \frac{H_L(s)}{1 - H_L(s) \sum_{i=1}^{n-1} H_i(s) \cdot k_i} = \frac{1}{R \left( 1 + \frac{L}{R \cdot n} s \right)} \]  

(21)

Since the contribution of each slave converter is supposed to be identical, we have:

\[ \sum_{i=1}^{n-1} H_i(s) \cdot k_i = (n-1) \cdot H_L(s) \cdot k_i \]  

(22)

and the solution is given by:

\[ H_i(s) = \frac{R}{n} \left( 1 + \frac{n+1}{n} \frac{L}{R} s \right) \quad n \geq 2 \]  

(23)

This implies that each slave converter has a static gain of \( R/n \) and presents a zero at \( \frac{n+1}{n} \frac{L}{R} \) sec.

The solution to our problem is therefore very simple and easy to realize. However, we must not forget that we have assumed perfect voltage sources which is not always the case. Also, in practice we know that a differentiator will amplify the existing noise in the system and therefore we must add a pole to limit the system gain at higher frequencies. For each application, it is necessary to quantify the effect of these imperfections as well as those due to parasitic elements of the magnet chains.

The results of simulation using second-order models as well as tests on four power converters have shown that the performance obtained is identical to that of a single power converter. For the converters of LEP, where the acceleration is relatively slow, the slaves only compensate for the static modes. The master performs the dynamic role for the entire load.
In this case the transfer function (20) becomes:

\[
\frac{I(s)}{V_c(s)} = \frac{n \cdot H_n}{R \left(1 + n \frac{L}{R} s\right)}
\]  (24)

where the gain and the time constant of the load is multiplied by the number of converters.

6. CONCLUSIONS

Analogue regulation systems provide a remarkably high performance in power converters at an extremely low price. The simplicity, coupled with the excellent quality of the components, gives a high level of reliability. For these reasons analogue compensation has a long future in front of it particularly for the new generation of power converters.

However, the application of digital techniques to the conventional type of power converter can give reasonable performance in many cases. Its use in the topologies we have described can bring elegant solutions particularly in the field of non-linear problems, reference forming etc, which are not simple to treat analogically. Further, the possibility of adaptive control is a very attractive feature of digital control.

However, when extreme high precision (10 ppm) and rapid response times (<100 ms) are required, we are confronted with the limits in speed and precision of analogue to digital converters today. The digital current transducer will be the basis of success in this domain.

REFERENCES

SWITCH-MODE POWER CONVERTERS

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ABSTRACT
This chapter covers the use of switch-mode power converters to power the electromagnets of an accelerator or collider with particular reference to their recent use in the LEP machine at CERN. Such converters are made up of several stages of conversion, the most important being the medium-frequency inverter. Each of these stages of conversion is described in detail with examples shown from the converters used for LEP. The various advantages and disadvantages of such systems are discussed as well as certain precautions to be followed in their utilization.

1. HISTORIC INTRODUCTION

1.1 Growth of higher-power switch-mode converters

From 1982 we followed the evolution of switch-mode techniques with great interest. It became apparent from contacts at various conferences and with industry, that the power available from switch-mode converters was increasing rapidly. This meant that they could possibly be of interest to our applications though many details required investigation.

1.2 The learning phase

After having established close contacts with the academic world as well as industry, a programme of prototype evaluation was launched. Where industry could provide ready made prototypes they were borrowed or purchased, where not, CERN constructed prototypes of its own. In short we looked at:

- a non-isolated 10 kW chopper for motor drives (Socapé/CH);
- a non-isolated 7 kW chopper for magnetic elevation (Thyssen/D);
- a resonant 20 kW converter with transformer (Van der Heem/NL);
- an 80 kW, 16 kHz, chopper constructed by CERN;
- a resonant 40 kW "Foch" converter constructed by Jeumont-Schneider (F) [1];
- an isolated (50 Hz transformer) 700 W, 16 kHz, H-bridge chopper (Servowatt/D);
- a phase controlled 700 W, 40 kHz synchronous rectifier constructed by CERN;
- a 700 W, 40 kHz, H-bridge chopper constructed by CERN.

At the end of this period we had learnt how to drive and use the electronic switches, how to reduce EMI to acceptable levels for microprocessor integration and p.p.m. loop performance, and how to design HF magnetic elements (transformers, chokes). We had also established:
- the dynamic and static performance advantages;
- the need for HF galvanic isolation for electrical noise reduction and for safety reasons;
- that up to 40 kW could be attained;
- which type of topology would be applicable;
- how to split the contracts.

2. **CONVENTIONAL LINE-COMMUTATED THYRISTOR CONVERTERS**

We will very briefly recall the structure of a 50 Hz conventional converter which has been presented elsewhere in these proceedings. It is shown in schematic form in Fig. 1 and consists of a 380 V (or higher), 50 Hz, network, an isolation and adaptation transformer operating at 50 Hz, a controllable thyristor bridge forming a voltage source, and a filter consisting of a choke and capacitor. Its characteristics can be summarised as follows:

- bandwidth limited to 150 Hz;
- energy reversal possible to the network;
- variable power factor from 0 to 0.8;
- harmonic content on input current which moves in phase;
- thyristor notches in network and noise spikes;
- large and heavy magnetic elements (transformer, choke);
- large residual ripple unless active filtering employed;
- simple structure, well understood;
- high power capability;
- moderate prices and competitive market.

![Schematic of a conventional thyristor line-commutated converter](image)

**Fig. 1** Schematic of a conventional thyristor line-commutated converter

3. **THE SWITCH-MODE TYPE OF POWER CONVERTER**

3.1 **Structure**

In this type of converter the line frequency is no longer the synchronizing and commutating element. The frequency of operation can now be chosen by the designer and, depending on the type of structure, the conducting and/or extinction moment can be chosen freely. The topologies of a switch-mode converter are normally more complex than those of the line-commutated thyristor converter. They can normally be split into three stages of conversion (see Fig. 2).
i) The input stage - Disconnecting element from 50 Hz network coupled with rectification and filtering to form a D.C. bus.

ii) The intermediate stage - Chopping in one way or another and modulation of an A.C. output (inverter or chopper).

iii) Output stage - Adaption to the load requirement by transformer if necessary, rectification and filtering.

Fig. 2 Structure of a switch-mode converter

3.2 Characteristics

Again their characteristics can be briefly listed as the following:

- high bandwidth, depending on the frequency used and response time;
- power factor of 0.95 if well designed. Form factor can be very bad however;
- energy recuperation to network not obvious;
- RFI doubts;
- small residual ripple at output;
- light weight, reduced volume;
- limited power capability;
- reliability questions;
- complex topology;
- new concepts, lack of specialists and suppliers.

4. INPUT STAGES

Either individually or on a common basis (say for all the converters in an equipment building) we need to supply a D.C. source for our inverters or choppers. For small powers, say below 500 W, this can be taken from one phase at 230 V giving a peak D.C. voltage of the order of 325 V. For higher powers and to help with single phase drops a 3-phase source is used giving a peak D.C. voltage of about 562 V.

4.1 Choice between individual and common input stages

This is not a simple choice, but for our application we have used individual input stages which give:

- complete independence and autonomy;
- simple protection and switching at A.C.;
- single source (3 x 400 V),
whereas the common input gives

- reduced cost since at larger powers the cost/kW is less;
- careful evaluation of power needs (extensions can be difficult);
- problems with circuit breaking of D.C.

4.2 The common input stage

There are many variations of the common input stage, some of which are shown in Fig. 3. In its simplest form a six-pulse diode bridge (I A) is connected to the network and feeds all units. Individual input filtering is achieved at each converter. An extension of this is to have one large filter (II A). However, decoupling capacitors are needed at the inputs of each converter to feed the pulsing nature of the following inverter. Stability on the D.C. bus can also be a problem because of the negative impedance effects of each unit.

With these systems, which would normally have the neutral to earth, a 150 Hz sawtooth of 78 V peak to peak is created which, if no galvanic isolation is provided in the converter itself, will form a common mode noise for the magnet. Power factors should be good (0.95) with harmonics of 5, 7, 11, 13 etc., which could be high for the supply system. The nominal D.C. voltage will be 565 V which might not be suitable for the switches of the inverters or the loads in the case of chopper-type supplies.

![Diagram](image_url)

Fig. 3 Variations of the common input stage
An alternative is to use a polyphase isolation transformer as shown in the two cases I B and II B. Several advantages now arise:

- we can earth the secondary where we like;
- if we phase-shift the secondary we can increase the system number to 12, thus eliminating the 5th and 7th harmonic on the input. Likewise ripple is small and, at 600 Hz, easier to filter;
- we can adapt the D.C. bus voltage to any level we desire.

With all these systems the stability of the D.C. bus has to be watched carefully and there is the problem of circuit breaking and/or isolation from the D.C. bus which is not as straightforward as for the A.C. case. Peak fault currents can also be very large depending on the size of the filter capacitors and impedance of the bus.

4.3 The individual input stage

The simplest input stage consists of an A.C. circuit breaker, RFI filter, diode bridge and pre-filter. While the choice of circuit breaker and diode bridge is relatively straightforward the filters have an important role. The RFI filter can normally be of a commercial type to reduce conducted noise into the network. Fast diodes are sometimes used for the rectifier. The pre-filter should normally have a resonant frequency in the range 100 to 200 Hz, its role being to reduce both the voltage spikes etc, coming from the network and the residual ripple. It also improves the input current form factor.

There are two other important factors to consider for the first filter. A regulated SMPS, since it works at unity power factor and takes a constant power, demonstrates negative impedance, unlike a thyristor line-commutated unit. As the network voltage rises, the current demanded by the inverter falls (IV = constant). This has the tendency to completely undamp the filter and cause oscillations (Fig. 4). The filter therefore should always assure positive resistance as seen by the network and be adequately damped itself (Fig. 5). One needs normally to apply critical damping to overcome this problem. It is also judicious to add a pre-charge circuit to minimize large switch-on currents.

![Fig. 4 Negative impedance effect on D.C. voltage variations](image1)

![Fig. 5 Schematic of the input stage](image2)
Many small-power, commercial designs use a simple diode and capacitor input stage drawing large peak currents. While this can be accepted for small quantities, it is totally unacceptable if large numbers are installed. In such cases an LC filter should be used, preferably associated with a three-phase rectifier.

4.4 **Galvanic isolation**

One can find many topologies which do not need galvanic isolation between the load and the supply network. However, somewhere along the conversion path, it is often useful if not essential for one of the following reasons:

- exact adaption of the voltage to a suitable level for the load thus avoiding the installation of more switching capability than is necessary;
- correct earthing of the D.C. side with the possibility to limit earth fault currents which can cause severe damage to the load;
- increased safety;
- reduction of noise on the D.C. side.

Further, if an HF transformer is used after an inverter, then no power will be fed to the load in the case of a switch failure. This can be an important feature for the well-being of the load which is often more costly and difficult to repair than the converter itself.

5. **INTERMEDIATE STAGE OR INVERTER STAGE**

5.1 **Structure**

The input is a voltage source which is non-reversible but bi-directional in current by virtue of the capacitors. (Energy cannot however be fed back to the network.) These capacitors are placed on the output of the input stage and also on the input of the intermediate stage very close to the switches. In the former case they can be electrolytic with suitable damping resistors but should be of polypropylene for the latter case in view of the high currents and frequencies needed for the inverter itself. The output of this stage will be alternating to drive a transformer/rectifier, of high frequency to minimize size and maximize response, a source of bi-directional current and, in general, reversible in voltage.

5.1.1 **Structure for low power (.300 W)**

Low power converters are used for television sets and in electronics in general to supply integrated circuits. They are most often of the flyback type using one bipolar or Mosfet switch to chop the current in the primary of a transformer. The transformer can have several secondaries so that +15 V, -15 V, and +5 V can be produced. Feedback is applied to one of the output voltages. Some integrated types of SMPS are limited to 48 V or 120 V on the input. Post regulation is often used. This entails the use of a linear regulator on each output to stabilize for load variations. Since input variations are taken care of by the feedback to the flyback converter, the voltage drop on these linear regulators can be kept low.
5.1.2 Bridge structure

All types of bridge arrangements are descended from the 'H' bridge. The four vertical branches contain four semiconductors switches while the horizontal branch contains the loads, LC circuits, primary of the transformers etc. The complete or symmetric bridge can be simplified to the asymmetric, the half bridge or the chopper. The different structures possible for a static converter have been enumerated [2] according to the type of source at the input and at the output, namely

- voltage or current source;
- bi-directional or not in current;
- reversible or not in voltage,

thus giving 64 cases, of which 48 are possible. If the input and output sources are of the same nature the structures are called "indirect". If not they are called "direct".

The switches are formed by semiconducting components and categorized (see Fig. 6) according to the commutation possibilities:

a) spontaneous commutation, the diode;
b) spontaneous at the opening, commanded at the closing, thyristor;
c) spontaneous at the closing, commanded at the opening, the dual thyristor;
d) commanded at the opening and closing, the transistor (bipolar, Mos, igbt) or the gate turn-off thyristor.

![Fig. 6 H topologies](image)

5.2 Operating principles

The function of an H bridge as an inverter depends very much on the source type, and secondly on the circuits found in the horizontal branch. Two distinct types of converter have evolved, namely the chopper or pure switch-mode converter and the resonant converter. The latter type uses a resonant circuit placed in the horizontal branch of the 'H', thus giving this type of converter its name. Several types of resonant circuit can be used, series, parallel, or series and parallel, the latter being called a "double-resonant" converter. Each one has a specific mode of operation as shown in Fig. 7.
The series-resonant converter operating below the resonant frequency is known as the "Schwartz" converter. The switches are normally asymmetric thyristors mounted in anti-parallel with diodes. The resonant cycle is started at the firing of the thyristor. The thyristor blocks at the passage of current through zero and the current commutates into the diode. Another "negative" cycle can be started at this moment or even before the diode has finished its conduction.

The series-resonant converter operating above the resonant frequency is often known as the Foch converter. The switches are dual thyristors mounted again in anti-parallel with diodes. Although dual thyristors do not as such exist they are normally constructed using a transistor and suitable logic. The switch closes spontaneously at zero voltage and the current starts from zero current. Switch opening is commanded and current is commutated to the opposite diode in the same vertical branch.

The double-resonant converter functions in the narrow band of frequency above the parallel resonance and below the series resonance. The switches are normally asymmetric thyristors or GTOs mounted in anti-parallel with diodes. The switch closing is commanded and if one chooses a discontinuous mode, the current will start from a value of zero. Blocking occurs at the zero current point.

5.3 Characteristics

5.3.1 Commutation limits of the switches

It is important to carefully define the commutation power in the plane I-U during the opening and closing of the switches. Switches have a defined "area of security" and large powers or fatalities may occur for commutation outside this area [3,4].

Hard commutation

If an analysis of the I and U curves reveals insupportable losses for the switch then it is necessary to reduce them by additional circuits to reduce the dV/dt at opening (RCD circuits) or to slow the dI/dt at closing (RLD circuits or saturable reactor).
In general the losses are transferred from the switches to the resistance of the commutation circuit. These losses increase with frequency. Elegant LCD (Knöll) circuits exist which are known as loss free.

In the case of a chopper, the recovery charge of rapid diodes can considerably increase the switch-on current of a switch when current is rapidly commutated out of the diode. This can cause strong emission at high frequency (0.1 to 5 MHz) which can become intolerable at high power levels (greater than a few kilowatts). The use of resonant converters greatly helps in this domain.

**Soft commutation**

The resonant converters employ “soft” commutation for their switches. In discontinuous mode the commutations are at zero current and the current has a near-sinusoidal appearance of relatively modest slope. For other modes of operation only one of the modes of commutation is forced. This requires a snubber circuit but this is normally rather simple. As well as obtaining low losses in the switches, the relatively gentle slopes of the current waveforms limit the losses in conductors and ferrites due to the absence of high harmonics [5].

### 5.3.2 Switch commands

The type of command given to the various switches alters the power fed to the load. Several types exist:

- **fixed frequency** modulation of the impulse width often known as Pulse Width Modulation (PWM). Variations exist such as the current controller with negative slope reference, and two dephased commands;

- **variable frequency** fixed closed-time again with variations such as one finds in resonant converters.

The switch commands can be open loop, i.e. acting directly on the inverter, or closed loop where the regulation can be slow by operating on the voltage of the output, semi-rapid by linear compensation of the absolute inverter current or rapid by non-linear compensation of the absolute inverter current. For the latter case the time between two successive commands for closing or opening each set of switches, can be varied rapidly over the entire range. These types of command are less common since they must be specifically adapted or invented for each topology. However, they will give the smallest delay times, and hence the greatest performance potential.

### 5.3.3 Regulation loops

The command of an inverter is driven by a closed loop which stabilizes the desired parameter, be it voltage or current. The stabilization is needed because of perturbations coming from the input source or from the load and normally need to function in the low frequency area (0 to 300 Hz).

In general the synthesis of a compensator is made all the easier the higher the bandwidth of the inverter. This would lead us to inverters working at about twice the frequency of the required bandwidth. In reality, because of the non-linearity encountered, together with safety requirements built into the command electronics, the bandwidth is often limited to 1 or 2 kHz for an inverter operating at 10 to 20 kHz.
5.3.4 Operating range

Most accelerator applications require a large output range from the power converters. This can vary from 10:1 to as much as 50:1 for certain applications. This imposition can often be very difficult for switch-mode power converters and requires a particular study of the snubber networks and the switches under these conditions. The inverters have a minimum switching time for either opening or closing, a period of recovery and a minimum frequency of operation within specification. The faster switching Mosfet is less troubled by such specifications, but powers are normally limited to a few kilowatts.

For the Schwarz type of resonant converter the minimum output depends on the degree of discontinuous operation, and therefore reduction in frequency, that can be tolerated before the output filtering becomes insufficient. For example, a typical 25 kW Schwarz converter gave 25 A output at 4 kHz and 125 A output at 17 kHz. In contrast for the Foch series-resonant converter the opposite is the case since frequency increases with diminishing output. The limit is therefore in the switching speed which can be high since the commutated currents are low. For example in a 40 kW converter we had 28 A at 65 kHz compared with 200 A at 19 kHz which corresponded to maximum output. A higher range can be achieved by using a system of two half-bridges controlled in an independent manner.

The double-resonant converters deliver minimum current when operating near the parallel resonant frequency given by the blocking circuit. Its value depends on the quality of this circuit and just how close one can approach its resonant frequency. We were able to obtain a range from 3.5 A to 150 A maximum, i.e. a range of 43:1 in current for a converter of 38 kW (Fig. 8).

![Impedance Plot](image)

**Fig. 8** Double resonant circuit impedance

5.4 Power and general limitations

Small switch-mode power converters of the chopper type now exist in large quantities and are used in applications from a few watts to ten or more kilowatts. Traction equipment now uses choppers of several hundreds of kilowatts mainly due to the performance of the GTO. Resonant converters operating at interestingly high frequency for accelerator utilization are now feasible up to about 100 kW.
The characteristics of the switches give the limit of utilization in voltage, current, power and speed. It is essential to respect a number of rules particular to each type of component (capacitors, conductors, etc.) which contribute finally to the limit of power and frequency of operation.

The factor Power \( \times \) Frequency can be used as a criterion on which to judge performance. However comparisons should be made at similar powers. Our figures were:

- F type: 67 MWh for 675 W (bipolar);
- D type: 190 MWh for 20 kW (high current converter);
- Prototype chopper: 320 MWh for 20 kW;
- C type: 430 MWh for 37.5 kW (half-bridge).

The feasible limit therefore seems to be around 1000 MWhz.

6. **OUTPUT STAGE**

6.1 **Structure and characteristics**

The input to an output stage is often a source of current either pseudo-sinusoidal or chopped. A small transformer (Fig. 9) is used to give electrical isolation and to adapt the impedance of the load to the inverter. The size of the transformer (or choke) is inversely proportional to the frequency of operation. One might therefore expect a 5 kHz transformer to be 1/100 of the size of a similar power 50 Hz version. Since the flux levels have to be kept lower, and losses due to high frequency are greater, this advantage is in practice reduced to about 1/20. Operating at such frequencies does however require close coupling and the use of special conductors. This imposes the same values, or integer values, of winding numbers on the primary and secondary. Voltage elevation or reduction is often obtained by the connection in series or parallel of the input or output of different windings on the same transformer or on several transformers.

![Fig. 9 A 50 Hz and a 12 kHz transformer of the same power](image)

Rectification is by rapid diodes either in a single phase full bridge, or by a half-bridge with centre point on the transformer; the voltage and current levels dictating the choice. Normally the output filter is made up of capacitors and inductors depending on the sources, and is small due to the higher frequency of operation. The first capacitor receives high pulsed currents and must therefore be polypropylene. A second stage of filtering is normally required to reduce noise to an acceptable level and electrolytic capacitors can often be used here.
6.2 Bi-directionality

In the simplest terms, if a smooth and uninterrupted passage through zero is not required, a polarity reversal switch must be used if bi-directionality of current is required. This can either be mechanical or electronic using thyristors. In this way decay to zero current of the magnet is assured by the rectifier diodes before changing polarity.

If true bipolar performance is required then a four quadrant solution must be sought. Absorption of the energy is a problem since the rectifier bridge of diodes is non-reversible. If this is not replaced by a synchronous rectifier then sufficient absorption must be provided in the capacitors of the output stage for the speed of operation required. Bipolar output stages of a four quadrant nature are very much more complex and, in particular, require energy dumps in the case of mains failure.

6.3 Performance and limitations

The residual ripple voltage at the output can easily be inferior to 1% and even attain 1/‰ of the maximum output voltage without too much difficulty. The sensitivity of the magnet diminishes with frequency at 20 dB/decade meaning that the high frequency residual ripple has little or no effect on the beam. In the case of high current power converters the output stage becomes important for the overall efficiency of the equipment. In these cases we use diodes of low forward voltage drop, the Schottky diode being an ideal candidate.

Power limits in this stage can normally match those of the inverter stage. At large current certain low voltage diodes can be put in parallel and one can also connect the transformers in parallel via the diodes.

7. ELECTRO-MAGNETIC COMPATIBILITY

All power converters must respect certain norms and switch-mode models are no exception. In fact having earned a bad reputation in this domain, the latter are often subjected to excessive restrictions. Norms such as VDE 875 or VDE 871 class A and C govern the radiated and conducted noise (Fig. 10). While conducted noise can be measured, we have always found it difficult to quantize radiated noise. Our converters are normally placed in metallic equipment buildings without windows and we therefore suffer little from this effect. We also ensure that the enclosures of the converters themselves are metallic.

![Fig. 10 EMI norms for conduction on input](image-url)
The conducted noise on the primary side of the converter comes from the input diode bridge and the inverter itself. The use of fast diodes and RFI filters on the input can normally resolve the former. Once again the use of soft commutation as found in resonant converters greatly reduces the noise produced by the inverter and/or allows higher power levels. It is essential to minimize the capacitance to earth of elements submitted to high dV/dt.

In general the level of noise conducted to the load depends on its susceptibility. The norms are usually less severe (often five times greater than those for the input) and for normal magnets can be achieved by the normal filtering. For high current supplies the problems can be more difficult. The layouts of the busbars need to be very symmetric and some additional filtering may be necessary when feeding superconducting magnets.

8. PRACTICAL CONSIDERATIONS OF SWITCH-MODE CONVERTERS

8.1 Technologies employed

The switches are evidently one of the most essential items in the switch-mode power converter world and are being continually improved. The designer now has a choice of several types of thyristor, diode and transistor from which to build his ideal switch. As another chapter [6] covers this subject in greater detail we will not go further here.

The drive circuits are very important to the well-being of the semiconductor switch, and it is important to follow manufacturers' recommendations in this domain. Strangely, whatever the type of switch, be it an asymmetric thyristor, GTO, bipolar or Mosfet, the drive waveform is often similar in the final analysis. They are characterized by a high input current to close the switch, followed by a plateau of voltage or current to maintain conduction, and finally a negative current to bring conduction to an end. Many of these drive circuits can be integrated with the switch if sufficient quantity is required.

If thermal isolation is needed, various thin strips, such as Mylar or glass, are used to isolate the semiconductor from the heatsink. More and more units are now being manufactured with the internal components isolated from the case. In this way the case can be bolted directly to the heatsink.

The most common cooling method is by natural or forced air cooling. It is efficient, cheap and reliable but can cause the converter to become coated with dust and grease after several years of service. In our case the buildings in which they are installed are relatively clean so this is not a problem. Other methods are available, such as mounting the heatsinks in the stream of forced air while keeping the components themselves in a relatively sealed enclosure. The use of heat-pipes can also be extremely attractive to take heat away to a remote area. In this way optimum cabling and minimized capacitance can be achieved. Alternatively, oil or water circulation can be used, or power devices can be located in sealed baths of Freon.

In general ferrites are used for the cores of the magnetic elements. They are becoming available in larger sizes and their field levels and operating temperatures are being increased. Ferrite transformers of 100 kVA are now possible. Insulated multi-strand wire or thin bands are normally used for conductors because of the skin effect. Likewise the transformer and chokes use "Litz" wire or bands. The dielectric of the capacitors is often polypropylene while thin aluminium bands are used for the conductors.
8.2 Switch-mode power converters built for the CERN LEP project

More than 600 switch-mode converters are now installed in LEP to power the Main Ring Magnet System. The horizontal and vertical corrector magnets are fed by 520 bipolar, four-quadrant power converters of 675 W output rating. The short chains of quadrupole, sextupole and mini-wiggler magnets with copper windings are powered by 108 modular resonant type power converters of a D.C. output rating of 37.5 kW. Eight high current 2000 A, 10 V converters feed the superconducting quadrupoles for the low-beta insertions in the physics interaction regions.

The converters have been designed for initial operation at 65 GeV with minimum cost, while being extendible to 100 GeV without waste. Switch-mode converters, with their reduced volume and modular approach, therefore seemed very attractive for the following three types:

<table>
<thead>
<tr>
<th>TYPE</th>
<th>POWER (kW)</th>
<th>U (V)</th>
<th>I (A)</th>
<th>QUANTITY</th>
</tr>
</thead>
<tbody>
<tr>
<td>C</td>
<td>37.5</td>
<td>125</td>
<td>300</td>
<td>108</td>
</tr>
<tr>
<td></td>
<td>188</td>
<td>200</td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>250</td>
<td>150</td>
<td></td>
<td></td>
</tr>
<tr>
<td>D</td>
<td>20</td>
<td>10</td>
<td>2000</td>
<td>8</td>
</tr>
<tr>
<td>F</td>
<td>675</td>
<td>± 135</td>
<td>± 5</td>
<td>520</td>
</tr>
</tbody>
</table>

The converters were purchased from European industry on the basis of precise performance specifications. While CERN suggested suitable topologies and block diagrams, the manufacturers were free to propose alternative solutions according to their own experience.

Converters of Type C

Using the same 37.5 kW inverter powered directly from the 380 V three-phase mains, type-C converters have three output modules as shown above. They are also capable of being connected in parallel to give 75 kW output. In this way all converters used for LEP 1 can be re-used in LEP 2. Alsthom's solution was to use a double resonant converter as shown in Fig. 11. Using GTO's in a zero-turnoff mode, the inverter works between a minimum frequency of 7 kHz, where it produces 10 W, and a maximum of 11 kHz for 37.5 kW. The desired power can be obtained from a half-H bridge. Four converters are located in one cubicle measuring 3.28 x 0.9 x 2 m as shown in Fig. 12 [7].

![Double-resonant converter topology of 37 kW](image)
Fig. 12 The modules of four double-resonant converters located in one cubicle

Converters of Type D

The series-resonant converter of the Foch type had for a long time seemed attractive, particularly because of its elegant commutation mode and since it suffered less from EMI problems. Again, qualifying prototypes were ordered from industry and tested with success, the contract for the series production being awarded to the firm Jema (E) whose expertise had been developed around an H-bridge inverter using Mosfet transistors. They therefore proposed the converter shown in Fig. 13. Since the designed bridge uses small Mosfets with limited voltage, an auto-transformer is necessary at the input so that the nominal D.C. rail voltage of the two inverters is 370 V. The inverters each feed three HF transformers connected in series on the primary, all secondaries being in parallel via low voltage Schottky diodes to the output filter. The inverters operate at 20 kHz and use PWM [8].

The power parts are air cooled by low speed fans. Efficiency at 2000 A, 10 V (the charging voltage) is 80%, while at 2000 A, 3 V (the steady-state voltage) it is 55%. Unlike thyristor line-commutated converters the power factor remains at 0.95 even at this low output voltage. The converter is located in a standard 19 inch rack, and is shown in Fig. 14.

Fig. 13 Topology of a high current switch-mode converter

Fig. 14 High current switch-mode converter modules installed in a rack
Converters of Type F

After competitive tendering for this type of converter, the contract was placed with GEC (GB). This converter consists of four stages as illustrated in Fig. 15:

- a rectifier with a six-diode bridge working directly off the three-phase 380 V mains;
- a half-bridge Mosfet inverter providing an isolation via an HF transformer;
- an HF rectifier with diodes;
- a switching inverter again consisting of Mosfets.

The last stage ensures the bi-directionality of the load current. It operates at 50 kHz and a 50:50 mark-to-space ratio gives 0 V output; by varying this ratio either positive or negative output can be achieved with a smooth transition through zero. The stored magnet energy can be transferred to a capacitor on the D.C. link, or dissipated in a power transient suppressor (a resistor controlled by a Mosfet). The view, Fig. 16, shows the two converter types with current ratings of 5 A and 2.5 A [9].

![Fig. 15 675 W converter block diagram](image)

![Fig. 16 Bipolar 675 W switch-mode converter](image)
9. **CONCLUSIONS**

9.1 **Comparisons between different types of converter**

One can compare the different converters by considering their power-to-weight, -volume or -surface ratios. Thus the table below can be established where the B type is a conventional thyristor unit, E a thyristor dual-converter, C a resonant converter, and D and F are chopper types.

<table>
<thead>
<tr>
<th>MODEL</th>
<th>B</th>
<th>13</th>
<th>25</th>
<th>9</th>
</tr>
</thead>
<tbody>
<tr>
<td>Power/weight W/kg</td>
<td>47</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Power/volume W/dm³</td>
<td>19</td>
<td>5</td>
<td>25</td>
<td>8.4</td>
</tr>
<tr>
<td>Power/surface 1) kW/m²</td>
<td>20</td>
<td>5.3</td>
<td>25</td>
<td>7.4</td>
</tr>
<tr>
<td>Weight/volume kg/m³</td>
<td>409</td>
<td>377</td>
<td>322</td>
<td>341</td>
</tr>
</tbody>
</table>

1) including access space around the converter

The efficiencies of switch-mode power converters and conventional units are much the same but can vary according to the application and power. Switch-mode in general will be much better than linear types of regulator and can also give improvements for large current outputs.

The price per kilowatt depends on the power of the equipment, the number of converters involved and the application. For this reason comparisons are difficult. However we have noted that in the medium power range (10 - 50 kW) the thyristor equipment can still be cheaper or, at worst, equal to that of a switch-mode resonant type converter. Economies for switch-mode come from the modular installation and the reduced space required. The switch-mode power converter is still a specialized product with a greater economic risk factor for the manufacturer coupled with higher design or development costs. On the other hand, the conventional converter has been fully developed for many years and is manufactured in a very competitive market. The fact that switch-mode can already be purchased at similar prices to the conventional types would lead us to believe that in the near future, with further development and a competitive market, they will become even cheaper.

Economies are achieved during installation of switch-mode converters due to their simple handling and reduced space requirement. Power distribution systems are in general smaller due to operation at unity power factor requiring lower installed kVA.

The improved performance of switch-mode power converters may or may not be an advantage for operation depending on the application. They give performance equivalent to that achieved by a conventional converter with an active filter but with full dynamic range. The modularity, light weight and ease of transport can considerably ease maintenance since repairs need no longer be carried out in situ while the accelerator waits.

9.2 **Standardization and system evolution**

One inverter type can serve several different types of output stage. Thus many variations of volts and amperes can be achieved based on a "standard" inverter for all applications of a similar power. The increase of power often required by an accelerator as it increases in energy can be achieved by adding further inverters in parallel. This is simple since they are normally current sources.
9.3 Future trends and needs

In the future the concept of a "standard" inverter with a kit of output stages will fulfil most of our needs [10]. In this way prices will be minimised and training of personnel facilitated. An increase in the power available from switch-mode converters is desirable so that they can become more widely applicable. Elegant bipolar topologies at high power and current should be investigated to meet the power requirements of orbit corrector magnets in the next generation of particle accelerators. Lastly, the dream of not only unity power factor but an input current containing no harmonics should be vigorously pursued. However it must be achieved at low cost and high reliability if it is to be a success in the accelerator environment.

* * *

REFERENCES


POWER CONVERTER SIMULATION

F. Bordry
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ABSTRACT
The major parts of a power converter simulation program are presented. Since matrix notation is the tool par excellence for describing the network problem in a form suitable for programming, the matrix approach to various aspects of dc, ac and transient analysis is described. Criteria for the choice of semiconductor modelling and the numerical method are discussed.

1. INTRODUCTION

As engineering systems become more complex, their design must rely on more sophisticated techniques. What is design? It is iterative analysis until certain specifications are met, and is characterized schematically in Fig. 1.

![Design process diagram](image)

Fig. 1 Design process diagram

Using the available technology (thyristor, transistor ...), techniques and experience, the engineer formulates a possible design to meet the specification. Normally he determines first the structure of the converter and then the ratings of the components. The proposed design must then be analyzed to find out if it is reasonable. The next step is to construct and test a prototype.

In recent years, simulation programs have played an important role in the analysis phase. These programs, which perform DC, AC, transient, and in some cases sensitivity and statistical analysis, are used in the design process by performing repeated analysis of a power electronic network. The major parts of a power converter simulation program are:

- modelling of the active and passive devices
- network topology
- equations determination
- numerical computation.
From the point of view of the end-user, these parts are hidden by user-friendly interfaces (pre- and postprocessor), but it is very important for the choice and efficient use of a simulation program to have full knowledge of these topics.

2. **MODELLING OF SEMICONDUCTORS**

To obtain meaningful results from a power converter analysis, passive and active linear and nonlinear devices must be represented to an acceptable degree of approximation. Furthermore, the model should be simple and present an interpretable equivalent circuit to the design engineer. It should also be accurate in describing the physical device to permit prediction of the circuit performance.

Several considerations must be taken into account when modelling an active device by an equivalent circuit:

a) **Operating region**: if the device performs over only a small range of voltages and currents, the parameters in the model may be considered as approximately linear. Otherwise the device model may require nonlinear parameter representation.

b) **Accuracy-Complexity**: generally, in any analysis, the degree of accuracy required will greatly affect the complexity of the equivalent circuit used in the modelling, and consequently the order of the equations (time cost).

c) **Measurability of parameters**: the numerical values for circuit elements in the model must be determined, but also their dependency on electrical and physical conditions. For a complete model, it is not very often possible to give all the parameters.

A power converter program is circuit-design (not device-design) oriented. Therefore the object is not to describe the physical processes of device operation, but to compute the circuit response.

We can divide the simulation programs into two classes, according to the modelling of the semiconductors:

- small-signal frequency domain
- switch-functioning

The first class of programs was developed for electronic problems. However, power-electronic engineers used them because they were the only ones available at the time. Recently, these programs have been developed for use in power-converter simulation.

Although there are many models for use with semiconductors, the following three have received much attention in modelling with transistors:

a) Ebers-Moll model
b) Beaupre-Sparkes charge-control model
c) Linvill lumped model.
Generally, the Ebers-Moll model is mostly used for analysis programs (Fig. 2). For other types of semiconductor, the models are derived from the transistor model. It should be pointed out, however, that the thyristor model is the stumbling-block for these kind of programs. Very often, there is simply no adequate model for a thyristor and this is a very serious limitation for the power-converter simulation.

\[ i_{re} = I_{re} (e^{\frac{V_{BE}}{AT}} - 1), \quad i_{re} = I_{re} (e^{\frac{V_{BC}}{AT}} - 1) \]

\[ C_{re} = \frac{C_D}{(V_B - V_{BE})^n}, \quad C_{re} = \frac{C_D}{(V_B - V_{BC})^n} \]

\[ C_{re} = \frac{q}{kT} \cdot i_{re} \]

\[ C_{re} = \frac{q}{kT} \cdot T_r \cdot i_{re} \]

Fig. 2 Ebers-Moll model

In the second class of programs, the semiconductors are considered as switching devices, i.e. nonlinear elements (they are either on or off). This assumption involves the representation of a semiconductor by a mono-output logical system (switch on or off). This binary representation leads to numerous models, the most frequently used being where:

- the semiconductors are considered as perfect switches. In the off-state, the branches on which they are situated disappear from the graph. The topology is thus variable and a particular system of equations corresponds to each sequence (varying topology).

- the semiconductors are modelled by a second-order circuit (inductance series and parallel RC circuit). The semiconductor itself is considered as a perfect switch; when the semiconductor is off, the variable state "current" associated with it is forced to be nil. The topology is fixed and the device can thus be described by a single system of equations (constant topology).

- the semiconductors are modelled by controlled voltage (or current) sources. The voltage at the bounds of a voltage source corresponding to a semiconductor in the on-state is nil, while for the off-state it is determined to produce no current in the source (dual proposition in the representation by a current source). The topology is fixed and there is a single system of equations, but at each instant the value of the controlled sources corresponding to the off semiconductors has to be calculated.

- the semiconductors are modelled in the form of a binary resistance, varying on a large scale according to whether they are off (high resistance) or on (low resistance). There is a single topology and a single state vector; the state equation coefficients depend on the value taken by the binary resistances. The semiconductor model may be improved, locally, by connecting in series or parallel any other components.
The drawback of this second class of models is the resolution of an extremely stiff differential equation (widely spread time constants). However, an ideal switch model (switch on: short circuit, switch off: open circuit) is used for each semiconductor, the stiffness problem is solved but the topology becomes varying and the switching-device voltage and current relations are difficult to formulate for computation. We shall come back to these problems further on.

3. **NETWORK ANALYSIS**

3.1 **Topological analysis (definitions)**

When two or more components are interconnected, the result is an electric network. Such networks store energy, dissipate energy, and transmit signals or energy from one point to another. A component of a network lying between two terminals to which connections can be made is called a branch. Two or more branches connected together are called a node. A simple closed path in a network is called a loop or a mesh.

The topological properties of a network are independent of the types of components that constitute the branches. So it is convenient to replace each network element by a simple line segment for the topological analysis. The resulting structure consists of nodes interconnected by line segments. There is a branch of mathematics, called the theory of linear graphs, that is concerned with the study of just such structures. A correspondence between a network and a linear graph can immediately be made. Thus the graph associated with the network of Fig. 3a is shown in Fig. 3b; each node and branch are indexed. We will now use this simple graph to illustrate some definitions or properties.

![Fig. 3 Correspondence between a network and its linear graph](image)

A graph whose branches are oriented is called an oriented graph. A way of storing the connectivity information is to define a matrix, called the incidence matrix, where the columns show which branches are connected to each node as well as the polarity of the connections (Fig. 4). Since each row of the incidence matrix contains both a +1 and a -1 as its non-zero entries, its columns are linearly dependent, i.e. their sum is zero so that any column is equal to the negative sum of all the other columns. Hence, any one column may be deleted since it contains redundant information. In our example, we choose to delete the column corresponding to node D in Fig. 3 and consider this node as the datum or ground node. The resulting matrix is called the branch-node matrix (Fig. 4b).
A subgraph is a subset of branches and nodes of a graph. A tree is a connected subgraph of a connected graph containing all the nodes of the graph but containing no loops. When specifying a tree, it is sufficient to list its branches. The branches of a tree are called twigs; those branches that are not on a tree are called links. Together they constitute the complement of the tree, called the cotree. The number of twigs on a tree is one less than the number of nodes of a graph.

If a graph is unconnected, the concept corresponding to a tree for a connected graph is called a forest, which is defined as a set of trees, one for each of the separate parts. If we choose a tree (a tree is not unique) in the graph and classify the remaining branches as links we may partition the branch-node matrix into submatrices referring to twigs and links ($A = A_T + A_L$).

It is easily shown that the submatrix $A_T$ is square and has an inverse which can be determined directly from the graph of the tree in terms of the node-to-datum path matrix $B_T$. For the tree shown in Fig. 5a, the columns of $B_T$ indicate which branches are included in the path from each node to the datum node. The relation between $B_T$ and the inverse of $A_T$ is:

$$A_T^{-1} = B_T^T$$

where $B_T^T$ is the transpose of the $B_T$ matrix.

---

Fig. 4 Incidence matrices

<table>
<thead>
<tr>
<th>Node</th>
<th>Branch</th>
<th>A</th>
<th>B</th>
<th>C</th>
<th>D</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>1</td>
<td>1</td>
<td>-1</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td></td>
<td>2</td>
<td>0</td>
<td>1</td>
<td>-1</td>
<td>0</td>
</tr>
<tr>
<td></td>
<td>3</td>
<td>0</td>
<td>0</td>
<td>1</td>
<td>-1</td>
</tr>
<tr>
<td></td>
<td>4</td>
<td>0</td>
<td>0</td>
<td>1</td>
<td>-1</td>
</tr>
<tr>
<td></td>
<td>5</td>
<td>-1</td>
<td>0</td>
<td>0</td>
<td>1</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>A</th>
<th>B</th>
<th>C</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1</td>
<td>-1</td>
</tr>
<tr>
<td>2</td>
<td>0</td>
<td>1</td>
</tr>
<tr>
<td>3</td>
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<td>0</td>
</tr>
<tr>
<td>4</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>5</td>
<td>-1</td>
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</tr>
</tbody>
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<table>
<thead>
<tr>
<th>A</th>
<th>B</th>
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<td>1</td>
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<tr>
<td>2</td>
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<td>0</td>
<td>1</td>
</tr>
<tr>
<td>5</td>
<td>0</td>
<td>1</td>
</tr>
</tbody>
</table>

Fig. 5 Tree with branches
The next topological matrix of interest for this presentation is the **branch-mesh** or **circuit matrix** (Fig. 6). Given a graph, first select a tree and remove all the links. Then replace each link in the graph one at a time. As each link is replaced, it will form a mesh or a loop. If it does not, it must have been a twig! This mesh will be characterized by the fact that all but one of the branches are twigs. Meshes (or loops) formed in this way are called **fundamental** (or **basic**) **meshes** (or **loops**). Consequently, the submatrix \( C_L \), referring to the links, is a unit matrix. Only the submatrix \( C_T \), which gives the path-in-tree for each fundamental mesh, is needed and it can be obtained as follows:

It is a standard theorem that, for any linear graph, the branch-node and branch-mesh matrices obey the fundamental relation:

\[
A^T \cdot C = 0 \quad \text{(or } C^T \cdot A = 0)\]

We may partition the matrices:

\[
\begin{bmatrix}
A_T^T & A_L^T \\
C_T & C_L
\end{bmatrix}
= A_T^T \cdot C_T + A_L^T = 0 \quad (C_L = I)
\]

This equation can then be solved using the relation between \( B_T \) and \( A_T \):

\[
C_T = - (A_T^T)^{-1} \cdot A_L^T = -B_T \cdot A_L^T
\]

The final topological matrix is the **cutset matrix**. A cutset is a set of branches of a connected graph whose removal causes the graph to become unconnected into exactly two connected subgraphs, and the two subgraphs contain all the nodes of the initial graph. Just as each link, together with a certain set of tree branches, defines a fundamental (or basic) mesh, so each tree branch together with a certain set of links defines a **fundamental** (or **basic**) **cutset**.

The columns of the fundamental cutset matrix (D, Fig. 7) show which branches are included in each cutset. With each tree branch positively oriented in its corresponding basic cutset, we obtain:
Thus the fundamental cutset matrix $D$ is computed in terms of $A$ and $B_T$.

### 3.2 Electrical network laws

We defined an electric network as an oriented linear graph with each branch associated with two functions of time $t$: the current $i(t)$ and the voltage $v(t)$. These functions are constrained by Kirchoff's laws and the branch relationships:

**Kirchoff's current law** (abbreviated by KCL) which states that in any electric network the sum of all currents leaving a node is zero, at each instant of time and for each node of the network. When this law is applied at a node in a network, an equation relating the branch currents will result. Attention must, of course, be given to the current references.

**Kirchoff's voltage law** (abbreviated by KVL) which states that in any electric network the sum, relative to the loop orientation, of the voltages of all branches on the loop is zero, at each instant of time and for each loop in the network.

**Branch relationships.** Consider as a general branch the configuration shown in Fig. 8. The branch is made of an impedance (or admittance) element $Z$ (or $Y$) in series with a voltage source $E$ and in parallel with a current source $I$. Thus, there are three distinct voltage and current variables to identify in this branch:

$J = i + I$

$V = e + E$

Ohm's law for this branch can be written in terms of the element voltage and element current variables $V$ and $J$. Thus we may express Ohm's law for the entire network as the matrix equations:

$V = Z \cdot J$

or

$J = Y \cdot V$
where the vectors $V$ and $J$ consist of the entire set of element voltage and element current variables for the network, and where the admittance matrix $Y$ is the inverse of the impedance matrix $Z$. The diagonal terms of $Z$ (or $Y$) are the self-impedances (or self-admittances) of each branch. The off-diagonal terms are the trans-impedances (or trans-admittances) between pairs of branches. If corresponding off-diagonal terms are equal, they correspond to mutual impedances or admittances. Active devices or dependent sources may be represented by off-diagonal terms in the $Z$ or $Y$ matrix.

![Fig. 8 General branch](image)

Combining Kirchoff's laws and branch relationships, we obtain the matrix relations:

$$C^t \cdot e = 0 \quad \text{and} \quad A^t \cdot i = 0$$

The topological matrix $C^t$ acts as an operator which sums all the branch voltages around each basic mesh while the matrix $A^t$ sums all the branch currents leaving each node.

If we introduce the vector $e'$ corresponding to the node-to-datum voltages and $i'$ corresponding to the mesh currents, we obtain:

$$e = A \cdot e'$$
$$i = C \cdot i'$$

We can summarize the basic interrelations between the network variables and the topological matrices by the diagram shown in Fig. 9.

![Fig. 9 Network variables and topological matrices](image)
3.3 **Electrical network problem**

The electrical network problem can be expressed as follows: "Given a network whose configuration determines the topological matrices A and C, its impedance matrix Z (or admittance matrix Y), the voltage source E and current source vector I, find the branch voltage e and currents i'.

There are four distinct ways to solve this problem:

1) **The mesh method**

From

\[ V = Z \cdot J \]

and

\[ J = i + I \]

\[ V = e + E \]

we obtain

\[ (E - Z \cdot I) + e = Z \cdot i \]

If we first multiply by \( C^t \) and take into account Kirchhoff's voltage law \((C^t \cdot e = 0)\), we find that:

\[ C^t \cdot (E - Z \cdot I) = C^t \cdot Z \cdot i \]

Introducing \( i' \) by:

\[ i = C \cdot i' \]

then

\[ C^t \cdot (E - Z \cdot I) = (C^t \cdot Z \cdot C) \cdot i' \]

where \((C^t \cdot Z \cdot C)\) is the mesh impedance matrix.

Thus we obtain \( i' \):

\[ i' = (C^t \cdot Z \cdot C)^{-1} \cdot C^t (E - Z \cdot I) \]

We can compute from \( i' \), the branch currents \( i (i = C \cdot i') \) and finally obtain the branch voltages

\[ e = Z \cdot i - (E - Z \cdot I) \]

2) **The node method**

Starting with the admittance form of Ohm's law, we can proceed to treat the network via the node method, following steps quite similar to the mesh method.

\[ J = Y \cdot V \]

\[ (I - Y \cdot E) + i = Y \cdot e \]

\[ (A^t \cdot I - Y \cdot E) = A^t \cdot Y \cdot e = (A^t \cdot Y \cdot A) \cdot e' \]

implies that \( e' = (A^t \cdot Y \cdot A)^{-1} \cdot A^t \cdot (I - Y \cdot e) \)
(\(A^t, Y, A\)) being the nodal admittance matrix.

Finally:
\[
e = A \cdot e' \\
i = Y \cdot e - (I - Y \cdot E)
\]

3) The cutset method

Instead of using \(e'\), we can use the set of tree-branch voltages \(e_T\) which are linearly independent. The branch voltages may be computed using KVL:
\[
e = D \cdot e_T \quad \text{(D: cutset matrix)}
\]

The KCL can be expressed with the D matrix:
\[
D_t \cdot i = 0
\]

Using these relations, we have:
\[
(I - Y \cdot E) + i = Y \cdot e \\
D_t \cdot (I - Y \cdot E) = D^t \cdot Y \cdot e + (D^t \cdot Y \cdot D) e_T
\]

which implies that
\[
e_T = (D^t \cdot Y \cdot D)^{-1} \cdot D^t \cdot (I - Y \cdot E)
\]

where \((D^t \cdot Y \cdot D)\) is the cutset admittance matrix.

The variables \(e\) and \(i\) are evaluated as follows:
\[
e = D \cdot e_T \\
i = Y \cdot e - (I - Y \cdot E)
\]

4) The mixed method (state-variable approach)

In addition to the three standard methods of analysis, there is a fourth which is beginning to be more widely used, especially to solve the transient analysis. We recall that twig voltages form a basic set of voltages in terms of which all branch voltages can be expressed. Similarly, link currents form a basic set of currents in terms of which all branch currents can be expressed. First we choose a tree. Instead of expressing the branch relationships as \(V = Z \cdot J\) or \(J = Y \cdot V\), we write mixed branch relationships as follows:
\[
\begin{bmatrix}
V_L \\
I_T
\end{bmatrix}
= 
\begin{bmatrix}
Z_L & H_{LT} \\
H_{TL} & Y_T
\end{bmatrix}
\begin{bmatrix}
I_L \\
V_T
\end{bmatrix}
\quad \text{(L: Link, T: Twig)}
\]

We apply now KVL and KCL. KVL applied to the fundamental meshes:
\[ C^T \cdot e = C^T_T \cdot e_T + e_L = 0 \] implies \[ e_L = -C^T_T \cdot e_T \]

and KCL to the fundamental cutset:

\[ D^T \cdot e = i_T + D^T_L \cdot i_L = 0 \]

implies that

\[ i_T = -D^T_L \cdot i_L \]

Combining the three last equations, we get:

\[
\begin{bmatrix}
Z_L & H \cdot e_T + C^T_T \\
H \cdot i_L + D^T_L \cdot Y_T &
\end{bmatrix}
\begin{bmatrix}
i_L \\
e_T
\end{bmatrix}
= \begin{bmatrix}
-i_L \\
i_T
\end{bmatrix}
\begin{bmatrix}
-Z_L & 0 & 1 & -H \cdot e_T \\
-H \cdot i_L & 1 & 0 & -Y_T
\end{bmatrix}
\begin{bmatrix}
i_L \\
i_T \\
i_L \\
i_T
\end{bmatrix}
\]

3.4 Methods of solving the network problem

Three basic methods are available for solving network problems, namely DC, transient and AC analysis.

3.4.1 DC analysis

The DC method permits DC or steady-state solutions of linear electrical networks to be obtained and often provides the worst case analysis and statistical analysis (Monte Carlo). The most straightforward way of solving DC network problems is by Gaussian elimination of the variables in the mesh or node equations:

Mesh method:

\[ P \cdot i = Y \]

Node method:

\[ Q \cdot e = X \]

with:

\[ P = C^T \cdot Z \] \hspace{1cm} \[ Y = C^T \cdot (E - Z \cdot I) \]
\[ Q = A^T \cdot Y \] \hspace{1cm} \[ X = A^T \cdot (I - Y \cdot E) \]

There are many numerical methods to solve these equations such as LU decomposition of P and Q with several refinements. However, a compromise must be made between accuracy and the amount of computational time.

The partial derivation or sensitivity of the nodal solution matrix with respect to any admittance (or impedance) change can be computed. We can differentiate the nodal equations:

\[ A^T \cdot (dI - Y \cdot dE - dY \cdot E) = A^T \cdot dY \cdot A \cdot e' + A^T \cdot Y \cdot A \cdot de' \]

or
\[ dc' = (A^1 \cdot Y \cdot A)^{-1} \cdot A^1 \cdot [dI - Y \cdot dE - dY \cdot (E + e)] \]

The last expression gives the variation in node voltages as a function not only of changes in admittances (dY) but also of changes in the I and E vectors. This equation is used in several programs for computing sensitivities. Solving nonlinear DC networks is not simple, the most general technique being the Newton-Raphson iteration method.

### 3.4.2 Transient analysis

Transient analysis provides the time-response solution of electrical networks subject to arbitrary user-specified driving functions. The transient problem presents the greatest computational difficulty since the numerical integration of the differential equation is time consuming for nonlinear networks and extremely slow when the network has a wide spread of time constant (see the switching model of a semiconductor).

One technique is to replace the integral-differential equations associated with a transient analysis by algebraic equations, developed at each time interval. A capacitance and inductance model allows this to be achieved as shown in Figs. 10 and 11. Thus, the time step has to be determined as a function of the time constant of the circuit. The accuracy and the stability of this method is really very dependent on this choice (e.g. ECAP program). Hence, the problem is reduced to a sequence of DC analyses.

![Fig. 10 Transient analysis capacitance model](image)

\[ J = \frac{C}{\Delta t} \frac{de}{dt} (t) \quad J(t) = \frac{e(t)}{\Delta t} - \frac{e(t - \Delta t)}{\Delta t} \]

![Fig. 11 Transient analysis inductance model](image)

\[ J(t - \Delta t) + \frac{\Delta t}{2L} e(t - \Delta t) \]

\[ J(t) = \frac{1}{L} \int e(t) dt \]

\[ J(t) = J(t - \Delta t) + \frac{\Delta t}{2L} e(t - \Delta t) + \frac{\Delta t}{2L} e(t) \]
State equation

Define a normal tree as a tree having as twigs all of the independent voltage sources, the maximum possible number of capacitors, the minimum possible number of inductors, and none of the independent current sources. Then we can partition the voltage and current vectors as follows:

\[
\begin{align*}
\mathbf{e} &= \begin{bmatrix} \mathbf{e}_T \\ \mathbf{e}_L \end{bmatrix} \Rightarrow \mathbf{e}_T &= \begin{bmatrix} E_T \\ \mathbf{E}_T \end{bmatrix} \quad \text{and} \quad \mathbf{e}_L &= \begin{bmatrix} E_L \\ \mathbf{E}_L \end{bmatrix} \\
\mathbf{i} &= \begin{bmatrix} \mathbf{i}_T \\ \mathbf{i}_L \end{bmatrix} \Rightarrow \mathbf{i}_T &= \begin{bmatrix} i_T \\ \mathbf{i}_T \end{bmatrix} \quad \text{and} \quad \mathbf{i}_L &= \begin{bmatrix} i_L \\ \mathbf{i}_L \end{bmatrix}
\end{align*}
\]

For the reactive elements:

\[
\begin{align*}
\mathbf{i}_c &= \frac{d}{dt} \begin{bmatrix} C_T & 0 \\ 0 & C_L \end{bmatrix} \mathbf{i}_c \\
\mathbf{i}_u &= \frac{d}{dt} \begin{bmatrix} L_T & L_T \\ L_L & L_L \end{bmatrix} \mathbf{i}_u
\end{align*}
\]

From these relations and the final equation of the mixed method, we can obtain the state equation for a linear network as follows:

\[
\frac{d\mathbf{x}}{dt} = \mathbf{A}\mathbf{x} + \mathbf{B}_1\mathbf{u} + \mathbf{B}_2 \frac{d\mathbf{u}}{dt} \quad \text{(State equation)}
\]

\[
\mathbf{y} = \mathbf{C}\mathbf{x} + \mathbf{D}_1\mathbf{u} + \mathbf{D}_2 \frac{d\mathbf{u}}{dt} \quad \text{(output equation)}
\]

where

\[
\begin{align*}
\mathbf{x} &= \begin{bmatrix} \mathbf{e}_c \\ \mathbf{i}_c \end{bmatrix} ; \quad \mathbf{u} &= \begin{bmatrix} \mathbf{e} \\ \mathbf{i} \end{bmatrix} \\
\text{State vector} & \quad \text{Input vector} \\
\text{Y output vector} & \quad \text{(any combination of the other variables)}
\end{align*}
\]

In the general case of a nonlinear network, we obtain:

\[
\frac{d\mathbf{x}(t)}{dt} = \mathbf{f}[\mathbf{x}(t), \mathbf{u}(t)]
\]

\[
\mathbf{y}(t) = \mathbf{g}[\mathbf{x}(t), \mathbf{u}(t)]
\]

where \( \mathbf{f} \) and \( \mathbf{g} \) are nonlinear functions of \( \mathbf{x}(t) \) and \( \mathbf{u}(t) \).
**Time-domain solutions of the state equations:**

The problem is then reduced to solving a first-order vector equation:

\[
\frac{dx}{dt}(t) = F[X(t), U(t)]; \quad X(t_0) = X_0
\]

The various numerical algorithms for solving first-order differential equations with an initial condition can be classified roughly into two groups, the so-called one-step and multistep methods. One-step methods permit calculation of \(X(t+k\Delta t)\) given the differential equation and information at \([t+(k-1)\Delta t]\) only. Multistep methods require, in addition, values of \(X[t+(k-1)\Delta t]\) and/or at other instants outside the integration interval under consideration \([t+(k-1)\Delta t, t+k\Delta t]\).

The most famous methods are:

**One-step:** Euler and Runge-Kutta methods

**Multistep:** Predictor-corrector methods such as Adams, Adams-Bashforth, Moulton, Milne, etc. and the Gear method (varying step).

The multistep methods require considerably less computation, compared with the one-step methods, to produce results of comparable accuracy. But, with multistep methods, it is rather difficult to change the step-size \(\Delta t\) once the calculation is under way and, in any case, they have to start with the one-step method. The Gear algorithm is specially designed to solve this problem, but it is not yet widely used in the simulation programs.

In the case of a linear system, we can solve the state equation using an exponential matrix:

\[
X(t + T) = e^{AT} \cdot X(t) + e^{AT} \cdot \int_0^T e^{-A\tau} \cdot B \cdot U(t + \tau) \cdot d\tau
\]

and there are many numerical methods for solving the problem of the \(e^{AT}\) computation. The calculation step \(T\) becomes an observation step because it is not linked to the smallest time constants of the circuit. This is not the case of the general integral methods, where the accuracy is very dependent on the integration step.

### 3.4.3 AC analysis (frequency domain)

AC analysis permits the steady-state solution to be obtained for linear electrical networks subject to sine-wave excitation at a fixed frequency. Since this analysis also allows automatic parameter modification, it is easy to obtain frequency and phase response solutions. Another approach is via the state equation of the network from which we obtain:

\[
\frac{dX}{dt} = AX + BU \Rightarrow s \cdot X = AX + BU
\]

\[
\Rightarrow x = (s \cdot I - A)^{-1} \cdot B \cdot U
\]

For the steady-state AC problems, the driving function can be written as:

\[
U(t) = U_0 \cdot e^{st}
\]
The steady-state response is then:

\[ X(t) = X_0 \ e^{st} \]

Thus:

\[ X_0 = (sI - A)^{-1} \cdot B \cdot U_0 \]

The problem is resumed to find an effective way of computing \((sI - A)^{-1}\). The most common method is to find the matrix \(X\) formed by the eigenvectors of \(A\):

\[ Ad = X^{-1} AX \]

where \(Ad\) is the diagonal matrix. The same \(X\) matrix also diagonalizes \((sI - A)\):

\[ (sI - A)_d = X^{-1} (sI - A) X \]

Thus:

\[ X_0 = X \cdot \left[ (sI - A)_d^{-1} \right] \cdot X^{-1} \cdot U_0 \]

Then the method requires only the inversion of the diagonal matrix \((sI - A)_d\) at each frequency.

3.5 Digital Computer Problem

Let us come back to the numerical determination and solution of the state equations for the transient analysis:

\[
\begin{align*}
\frac{dx}{dt} &= F[X(t), U(t)] \\
Y &= G[X(t), U(t)]
\end{align*}
\]

If these equations describe the behaviour of a linear time-invariant network, their solution is straightforward. Using numerical integration algorithms (mono- or multistep method) or matrix exponential \([\exp(\Delta T)]\), the state equation is solved sequentially at discrete time intervals \(\Delta t\). At each step, the output equation is also solved but, being an algebraic equation, its solution is easily obtained.

The choice of \(\Delta T\) for the integration algorithms is critical: the larger the eigenvalues of the state matrix, the smaller \(\Delta t\) must be to assure convergence and accuracy. The drawback of choosing very small values of \(\Delta t\) is, of course, computation time.

For power electronic networks, the solution of the state equation is not nearly so straightforward. As previously described, switching changes the state matrices and, according to the switching models used, we have constant or varying topology.
3.5.1 Constant topology

The constant topology simulation is conceptually simple. Switching devices are modelled by time-varying impedance: high impedance (off), low impedance (on). The result is a fixed state equation whose coefficients vary with time. Only a single circuit analysis is needed to determine the state equation for all simulation time. At each step, the state of each switching device must be checked. A backward iteration may be necessary with a reduced step in the case of spontaneous commutation (diode, thyristors turning off ...) to verify threshold conditions.

The use of time-dependent impedances to represent the switching devices introduces a wide spread of time constants. To insure a converging and accurate solution, the integration step length of the state equation must be on the order of the smallest time constant. This fact leads to numerical instability (\(\Delta t\) too large) or to unduly time consuming computations.

Let us present a very simple example for a chopper circuit (Fig. 12). The spread of the time constants is in the order of \(10^5\). In a more complex system, the ratio can be in the order of \(10^9\) or \(10^{10}\). This point is fundamental in the choice of the numerical method though it is not obvious for all commercial programs which often have a fixed step. Gear's algorithm is still the most efficient method for nonlinear systems (multistep and varying-step method), and for finding the matrix exponential of linear systems (the computational step is an observational one).

\[
L \frac{di}{dt} = \left( R + \frac{r_d}{r_d + r_T} \right) i + \frac{r_D}{r_D + r_T} E
\]

\[
\text{Time constant: } \tau = \frac{L}{R + \frac{r_d}{r_d + r_T}}
\]

ON state: \(r_D = r_T = 0.1 \Omega\) \[T_{ON}, D_{GF}: Z_{GF} = 10^{-4}\]

OFF state: \(r_D = r_T = 10^6 \Omega\) \[T_{OFF}, D_{ON}: Z_D = 10^{-4}\]

\[T_{OFF}, D_{GF}: Z_{GF} = 10^{-9}\]

Fig. 12 Solution for a chopper circuit
3.5.2 Varying topology

If the switching devices are actually represented as switches, i.e. zero current when off and zero voltage when on, then the network will have a time-varying structure. Each state is then represented by a distinct state equation. When a state change occurs, the network incidence matrix is changed and with a series of matrix manipulations the state equation may be obtained.

These methods are very expensive in computation time. Commercial programs using varying topology are not available though some academic programs do exist.

An important point which has not been mentioned above, is the influence of the network complexity on the computation time. For instance, the simple addition of snubbers (RC circuit) to the circuit of one rectifier (Graetz bridge) could increase the computation time (and cost) by two orders of magnitude. The semiconductor model is, of course, very sensitive in this respect and the computation time can be very high especially with the Ebers-Moll and resistor model. It should also be remembered that the aim of these two models is very different; the former is for device study, the latter for system study.

4. AVAILABLE CODES

The available commercial or academic codes may be classified in three families as follows:

1) **Electronic network code**

   The most used are: ECAP, SPICE and derivatives Pspice, MicroCap, SCEPTRE ...

2) **Power-electronic network code**

   Commercial products are not available, presumably because the market is too small. However, some large industrial groups or university laboratories have developed their own programs and provide programming support to users; e.g.: NETASIM (AEG), KOPL (Siemens), HC100 (BBC) or NETCAP (University K. Japan), ATOSOE (Université Trois-Rivieres, Canada), SCRIPT (Université Toulouse, LEEI).

3) **Analog system simulators**

   The system simulators are designed to solve all systems that are definable using piecewise-continuous algebraic and differential equations. The main focus is generally electrical but, using a modelling language, other domains such as chemical, mechanical, hydraulic, can be explored. Unfortunately, the price and training costs for these simulators are not negligible.

   Often, in the complete design process, the three kinds of programs are required. As an example at CERN, Micro-Cap or Spice (Pspice), SCRIPT codes and SABER simulator are used by power electronic specialists.
5. APPLICATION EXAMPLE

As an application, a simulation (by SCRIPT program) of a high-voltage power converter is presented (Fig. 13). This 100 kV power converter (4 MW) feeds two klystrons in parallel (CERN Large Electron-Positron accelerator storage ring LEP). The thyristor ac controller is placed in the lines feeding the step-up transformer (TR3/4) whose secondaries feed two full-wave diode bridges. A symmetrical LC filter then provides smooth dc output at 100 kV, 40 A maximum.

![Electrical circuit of a thyristor controlled dc power converter rated at 100kV, 40 A.](image)

Fig. 13 Electrical circuit of a thyristor controlled dc power converter rated at 100kV, 40 A.

The klystrons are protected by a fast crowbar. In the case of a klystron fault, which may happen several times per day, the crowbar shorts the HV output terminal to ground. Obviously, the power supply has to be switched off rapidly and without damage under these severe conditions.

The waveforms of steady-state conditions at nominal point are given in Fig. 14 (the firing angle is 55°).

In order to simulate the converter behaviour under klystron crowbar-protection conditions, the output resistor (modelling the klystron) is shorted with a switch. Figure 15 shows this transient condition shortly after steady-state has been obtained. A start-up transient is also presented.

This simulation gives realistic component ratings for all components of the converter but in particular for those of the output diodes and filter.

6. CONCLUSION

In recent years, a great deal of attention has been given by various companies to improving the interfaces (schematic input, user-friendly post-processor ...) used in simulator programs. From the topics discussed in this paper, it is evident that work is needed to upgrade the switching device models and the numerical computation method for power electronic networks. In the near future, these two points will receive great attention in order to drastically improve the efficiency of power converter simulation programs.
Fig. 14  Steady state waveforms for the HV power converter

1) Voltage across ac controller
2) ac line-to-line voltage across secondary of HV transformer
3) ac line-to-line voltage across primary of HV transformer
4) ac winding current, secondary of HV transformer
5) Current through ac controller
6) dc voltage across one arm of HV diode rectifier

Fig. 15  Start-up and short-circuit transients

1) Current through ac controller
2) Current through one arm of HV diode rectifier
3) Current through filter choke
4) Voltage across filter choke
5) Output voltage
6) dc voltage across HV diode rectifier
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THE VARIOUS SYSTEM ASPECTS OF THE MAIN POWER SUPPLY OF THE CERN - SPS

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ABSTRACT
The main power supply of the CERN - SPS was built more than 15 years ago. Its performance, control and interlocks were continuously improved in order to meet the constant evolution of the machine requirements. It was originally designed to accelerate protons from 10 to 400 GeV/c, then improved to reach 450 GeV/c. A few years later, the SPS was adapted as a storage ring to generate collisions between protons and antiprotons. Recently, the SPS has been used as the LEP - Injector, accelerating electrons and positrons from 3.5 GeV/c to 20 GeV/c. After a global presentation of the main power system (for the dipole and quadrupole magnets), the experience acquired in the power part, the controls, the interlocks, the difficulties encountered and the reliability are presented in this paper.

1. INTRODUCTION

A description of the load (dipoles and quadrupoles F + D), which is located in the SPS tunnel (circumference of about 7 km), and all the different required cycles is first given. The design of the individual power converters, with some comments about the choice of the solutions, and their connections to the magnets are then presented.

After this presentation of the power part, we will then describe and comment in the next section on the evolution of all the different control systems which were used till now (field and voltage, current and voltage, current control with one and two loops).

In the last part, we will communicate some of the experience accumulated over years related to the operational aspects. Fulfilling the required accuracies and performance of such an equipment is fundamental, but getting it to run with a high reliability is certainly also very important. So we will describe our system philosophy related to the interlocks, diagnosis and fault finding principles, and which aims to minimize the down time of the equipment. We will also present some major difficulties encountered, especially during the first years of operation of the SPS. The final aim of this lecture is to transmit to you a part of our experience with such equipment, which is widely spread over the site, while trying to give to you some advice.

2. DESCRIPTION OF THE LOAD AND THE VARIOUS CYCLES

2.1 Description of the load

The complete load is distributed all around the tunnel. It consists of 744 dipole magnets and 228 quadrupole magnets (half of them being focusing quadrupoles, the other half defocusing).

Dipole circuit: All dipoles are connected in series via two non-isolated watercooled pipes in which the current flows in opposite directions in order to minimize the stray field due to this current and to reduce the inductance of the connections. As the voltage against earth was limited to 4 kV peak at the starting
date of the SPS, (then increased to 6 kV), we were obviously forced to insert power supplies all around this load. Twelve converters were initially distributed along this load but were later increased to 14.

**Quadrupole circuits**: One quadrupole out of two is a focusing one, the other being defocusing. One circuit connects all focusing quadrupoles in series, and the other, the defocusing ones. The two currents have about the same intensity and flow in opposite directions. Only one power supply feeds each system of quadrupoles. The connections are in fact rather more complicated, as we will see later, but it does not affect the main parameters of the load.

Reference magnets are located in an auxiliary building. In Table 1 the quadrupole parameters are given per individual circuit.

<table>
<thead>
<tr>
<th></th>
<th>Dipoles</th>
<th>Quadrupoles</th>
</tr>
</thead>
<tbody>
<tr>
<td>Resistance</td>
<td>3.25</td>
<td>1.25</td>
</tr>
<tr>
<td>Inductance (non sat.)</td>
<td>6.6</td>
<td>2.0</td>
</tr>
<tr>
<td>I min</td>
<td>38</td>
<td>14</td>
</tr>
<tr>
<td>I peak</td>
<td>5754</td>
<td>2200</td>
</tr>
<tr>
<td>dl/dt peak</td>
<td>1900</td>
<td>800</td>
</tr>
<tr>
<td>U peak</td>
<td>25000</td>
<td>3700</td>
</tr>
<tr>
<td>U flat-top</td>
<td>18700</td>
<td>2750</td>
</tr>
</tbody>
</table>

**2.2. Description of the cycles**

Three different pulsed cycles are used:

The **fixed target cycle**, (actual duration of 14.4 s) to accelerate mainly protons from 14 to 450 GeV/c, providing fast, fast-slow and slow extractions during the flat-top.

The so-called **lepton cycle**, (duration 1.2s) to accelerate leptons from 3.5 GeV/c to 20 GeV/c in order to fill the LEP - collider.

The **collider cycle**, used to accelerate simultaneously protons and antiprotons from 26 to 315 GeV/c, during the preparation of SPS collider coasting beams (315 GeV/c being the final coast level of the collider).

The fixed target cycle can be followed by four short lepton cycles during the necessary dead time imposed for thermal reasons. This combination is called "**interleaved cycles**" and allows LEP to be filled without affecting at all the physics done during the fixed target cycle.

**2.2.1 Fixed target cycle (14 - 450 GeV/c)**

As can be seen in Fig. 1, this cycle is composed mainly of the following parts:
flat bottom  1.26 s
acceleration  3.48 s
flat top      2.58 s
descent      1.62 s  duration: 8.94 s

It should be noted that the peak power for the dipole magnets, at the end of acceleration, is 145 MW, while it decreases to 107 MW during the flat top. If the reactive power during the rise is essentially due to the commutation of the converters, it jumps up to a high value during the flat top. This forces us to have a local compensation of the reactive power in order to avoid too large voltage fluctuations. The allowed mean power imposed by thermal cooling of the magnets is 32 MW, which leaves a dead time of 5.28 s for the lepton cycles, the power needed for the lepton cycles being negligible.

2.2.2 Lepton cycle (3.5 - 20 GeV/c)

The current at the flat bottom is very low: 38 A or 0.7 % of the peak current, which means running at the static limit of the interrupted current. The current then rises in 360 ms to 230 A with a maximum slope of more than 1 kA/s to reach the short extraction level of 238 A, followed by a second rise to 272 A and then the descent to the injection level. The second rise (without beam) exists only in order to ensure the same magnetisation of the magnets, independently of adjustments of the extraction level.

2.2.3 Interleaved cycles

The oscillogram in Fig. 3 shows the final combination of the fixed target cycle and the four Lepton cycles, with the same scale. Using the same power part for the two types of cycles requires different control loops to achieve the necessary accuracy.

2.2.4 Collider cycles

These are very similar to the fixed target cycles, but stop at the upper thermal limit of 315 GeV/c, which is the beam-coast level. A special control loop is adapted to the required high stability. We will not describe further this cycle, which is not interesting in this context.

3. DESCRIPTION OF THE POWER SUPPLIES

3.1 Individual converter

As shown in Fig. 4, each dipole power supply is composed of two sets of 12-pulse thyristors connected in series, with two devices in parallel. Each one is supplied from the 18 kV, 50 Hz, central busbars by a double transformer delivering the secondary voltages with a phaseshift of ± 15° electrical. A passive filter of second order, associated with a damping path, ensures the reduction of the output ripple of the converter. A 100-cycle tuned filter prevents subharmonics caused by any mains asymmetry. A further decrease in the final ripple is achieved by injecting a small ac current delivered by a transistor active filter at the series-choke terminals. The two 12-pulse sets are controlled in a sequential mode (one being in full rectifier mode, while the other is in full inverter mode to obtain zero voltage, for instance), in order to minimize the reactive power. The power supplies are completely floating, the two output terminals being just connected through small capacitors to earth. With the exception of the transformers, all power parts are water-cooled.
Protection devices placed on the converter output consist of three spark gaps; two connected to earth in order to protect the dipole magnets, and one (calibrated to a lower value) to avoid any overvoltage which could accidentally occur if one power supply failed to fire when the others (connected in series) just start to deliver voltage. It could also help in the case of an unwanted oscillation of the power supply near the resonant frequency of the passive filter. Any firing of a spark gap triggers the firing of the by-pass thyristors, the spark gaps being unable to dissipate energy without burning. As we will see later, these by-passes are also fired to discharge the energy stored in the dipoles (110 MJ peak) in case of faults. Series resistors limit the current if the by-pass thyristors are accidently in the short-circuit condition. Finally, a remotely-controlled bipolar switch allows the power supply to be connected to, or to by-pass, the dipole circuit. A neutral position is also provided to facilitate the location of faults on the load part.

The layout of the quadrupole power supplies is shown in Fig. 5. There is no need for thyristors in parallel, but there is one filter per 12-pulse set, in order to produce similar voltages and currents in each half of the focusing quadrupole load. This allows the simultaneous extraction of the protons during the flat top, by having 9 integer horizontal beam oscillations over one third of the machine, while the total number is about 26.6 around the whole machine.

We will now discuss some fundamental choices made in the construction of these power supplies:

- **18 kV circuit breakers.** These are all centred in the electrical building. They are relatively fragile and it is advisable to minimize their frequency of operation.

- **Rectifier transformers.** Accurate phase shift of 30° electrical between the two transformers of a 12-pulse set, and the equality of the short circuit impedances are the two main requirements for such devices. This can only be achieved by using a scheme with elongated deltas on the primary windings and a phase shift of ±15°. The use of a screen between the primary and secondary windings is not recommended. Any solution using a Wye-Delta coupling would produce worse results.

- **Thyristors.** Correct voltage rating is not a problem using a safety factor between 2.7 and 3. With two 12-pulse sets in series, each thyristor must be triggered eight times per period in order to ensure a correct path for the current, especially at low interrupted current. The current sharing for the two thyristors working in parallel must be achieved within a tolerance of the order of ±20% of the peak current. Then comes the question: must we use fuses or not? In the case of two thyristors in parallel, no selectivity and no emergency-service can be ensured with fuses. Additional safety margins must be provided in the case of pulsed current. This is expensive, requires more room than the thyristors, demands the use of interlocks, and incurs non-negligible losses. For all these reasons, it was decided not to use fuses. This means that a faulty thyristor will burn (40 kA symmetrical short-circuit current) and will probably involve the other thyristors of the same star which must not explode. All the rest of the material must withstand the short-circuit without damage.

- **Current sharing.** As was mentioned previously, it is not possible to connect two thyristors directly in parallel. The thyristors are assembled as complete three-phase bridges (six thyristors), supplied from the transformer by individual cables (four times three monopolar cables of 12 m length per bridge). The mutual inductance between cables from one bridge to the other must be minimized and this is achieved by optimum mixing of the phase. During the overlapping, the slope of the current is identical and the quasi-
static current sharing is ensured by the resistive voltage drop in the cable which is about the same as the voltage drop at the thyristor. This principle works perfectly well and has never caused any trouble.

3.2 Overall connections

The rectifiers are located in six equipment buildings, located around the machine. They are fed via individual circuit breakers by two separate 18 kV busbars located in the electrical building. Cables of various length, but a maximum of 3 km, bring the power to the individual converters. To facilitate repairs and maintenance, remotely controlled isolators permit the 18 kV input to the transformers in each equipment building to be isolated and earthed locally, without someone having to go to the building concerned.

As shown in Fig. 6, the dipole magnets are connected in series to the output of the 14 power supplies in order to allow good equalization of the voltage to earth. The current circulates in two opposite loops to avoid induced flux. In the tunnel, these connections consist of non-isolated copper pipes located behind the magnets and near the floor. They are supported by insulated brackets and are protected by a metallic cover.

Usually, operation is ensured by 12 dipole power converters, two being by-passed, as spares. If possible, a regular distribution of the converters, is used all around the machine. Figures 7 and 8 show respectively the voltage distribution to earth in a normal case, and with the emergency use of a spare power supply (calculated at the peak voltage, end of rise).

Figure 9 shows the connections of the quadrupole power supplies. Two points should be noticed. Firstly, there is a midpoint connection to the focusing quadrupoles in order to control separately, and with only a small difference, the current in the two halves of the machine. This is required in order to have a complete number of horizontal beam oscillations between the two extraction points, which are used simultaneously, with a non-integer number of oscillations over one complete turn of the machine. Secondly, there is a spare power supply which can replace either the focusing or the defocusing one.

As for the dipoles, the power supplies are connected to these quadrupoles via water-cooled, non-insulated, copper pipes.

3.3 Reactive power compensation

The fixed target cycle requires not only a large fluctuating active power, but also provokes large variations of the reactive power due to the control of the thyristors. The active power can only be delivered from the mains, while the reactive power, which would cause large voltage fluctuations, can be compensated locally on the 18 kV busbars. Each busbar is fitted with a reactive power compensator. In our case, as shown in Fig. 10, these use a saturable reactor, which acts as variable inductive load and a capacitive load shared into various filters, tuned from the 2nd to the 17th harmonic. The presence of the small chokes does not affect the capacitive performance at 50 Hz. This compensator is able to maintain the 18 kV voltage within the limits of ± 1%. The total capacitive power of each compensator is of the order of 100 MVar, while the possible excursion reaches about 70 MVar. During the flat top, there is about the same amount of active and reactive power (120 MW and 120 MVar for dipoles and quadrupoles).

The various filters short circuit the harmonic currents generated by the rectifiers and avoid distortion of the voltage waveform at 50 Hz. These two compensators work perfectly well, as long as the three phases of the
incoming mains are symmetrical. However, in the case of an unsymmetrical perturbation, they can generate harmonics, as described in section 5.3. When a compensator is switched on, it can also cause fluctuations for one to two periods. For this reason, any sensitive equipment, such as bidirectional rectifiers for instance, must be switched off before the compensator is switched on.

4. CONTROL SYSTEMS

The principle of the various control systems which were used to drive the main rectifiers of the SPS, from the beginning in 1976 until recently, will now be described. In all cases, a rapid and accurate voltage regulation is needed for each rectifier, with a good rejection of the perturbations coming from the mains or from the common-mode voltage of the dipole rectifiers. This was achieved by means of our standard analog gate and fast voltage control system.

4.1 Field and voltage control

This first control was based on field regulation. The basic idea is perfect being based on the measurement, in two dipole reference magnets placed outside the machine, of the true integral of the field seen by the beam over its trajectory. For the quadrupoles, the same principle was used. The tracking was done by reference to the computer program.

4.1.1 Principle

As drawn very schematically in Fig. 11, the field was measured and the value stored every 30 ms over the whole running cycle. The error resulting from the difference between the required field and the measured one was then used to generate the voltage reference, with the appropriate algorithm for the next cycle. This method offered a reliable measurement device and supposed that the load parameters do not change from one cycle to the next one (except if the machine starts from a cold state).

4.1.2 Measurement

This was achieved by placing a long coil in the stainless steel vacuum chamber of the two reference magnets of the SPS. The coil output voltage was applied to a voltage to frequency converter (VFC) delivering one pulse every $2.10^{-6}$ T. The frequency reached more than 500 kHz during the rise and was not far from MHz during the descent. At the end of the descent, a "peaking strip" was used to accurately preset the field value at the end of each cycle. This device was made with a magnetic wire placed in a well stabilized compensation field, in the middle of the reference magnet. When the main field passed the level of the compensation field, it induced a small pulse, triggering at this moment the preset value in the field counter.

4.1.3 Difficulties

As shown in the Fig. 12, the chosen VFC could work only with positive voltage, so requiring two converters (up and down) or an inverter. Both solutions are difficult to implement since it is almost impossible to find two identical converters. Using an auxiliary inverter, a switch is needed. Any asymmetrical ripple, especially during a flat part of the cycle, could not be integrated correctly. The exact triggering produced by the peaking-strip was influenced by the slope and the ripple which could vary at the end of the descent. This caused a certain jitter in the preset pulse. There were sometimes parasitic pulses which provoked unwanted errors as
shown in Fig. 14. Fault finding was rendered difficult due to the fact that these errors could appear anywhere in
the measurement, the calculation or the digital transmission of the voltage reference to the rectifiers. Due to the
wrong integration of the ripple and the non-ideal offset, it was not possible to use this method for long flat-tops or
for collider runs.

The SPS ran for several years with this type of control and was continuously improved. However, it
was very difficult to reach a stability better than $10^{-4}$.

4.2 Current and voltage control

The second type of regulation was a mixture of voltage and current control, as illustrated in Fig. 15.

4.2.1 Principle

Knowing the field tables, we can produce a program to pre-calculate the voltage tables needed in order to
deliver the expected current, provided that the magnetisation curve of the magnets is well known. The voltage
references are generated every 30 ms and sent to (n-1) rectifiers where n depends on the number of rectifiers left
as spare. One, but only one, rectifier will then be used as the current regulator, like a normal rectifier doing its
own current regulation. If all is well calculated, this rectifier should always work around zero voltage. This
implies a knowledge of the lag of the current regulator, in order to anticipate the current reference by the correct
advance.

4.2.2 Results

A very good reproducibility was achieved by this method. It was however very difficult to establish the
correct magnetisation curves for the dipoles and the quadrupoles at the beginning. With this principle, one
rectifier is "lost" in voltage and all rectifiers are no longer interchangeable. After a difficult changeover from the
field control to the current control, due especially to the uncertainty of the magnetisation curve, the voltage
calculation and the introduction of the voltage limitation, the situation was much more stable and repetitive than
before.

Two weak points were incurred: errors introduced by the digital transmission of the voltage references
to all the rectifiers, which can introduce errors and the limitation of the overall gain of the current loop (equivalent
to one rectifier supplying the total passive load of dipole magnets).

This kind of loop has also been used successfully for several years.

4.3 Current control

In order to have more gain in the current loop, it was evident that we should drive all dipole rectifiers
with the same analog control signal. The only problem was to test how an analog signal could be sent over
distances of 3 km without noise problems, and whether the different transmission delays could cause difficulties.
Tests made with long cables showed that it might be possible to achieve a simple, though not perfect,
transmission. So we tested the solution presented in Fig. 16, and the results were very positive.
4.3.1 Principle

This is a simple cascade regulation current-voltage method, all rectifiers being driven by the same analog signal. Many tests were made to check this simple transmission system and to look at noise effects.

4.3.2 Results

The new control system has been in use for about one year now. The maximum current gain can be obtained, as if one big rectifier were supplying the whole load. The drawbacks of the transmission system are very few, though improvements are still needed, especially in relation to noise. Its big advantage is its simplicity. Another advantage is to have identical rectifiers all around the machine. They are all used within the possible voltage. So, two rectifiers can always be left as spare, all possible configurations with 12 rectifiers being operational. The suppression of digital signals sent continuously over long distance seems to be advantageous.

4.4 Current control with two current loops

Another improvement, a double current loop, has recently been installed (see Fig. 17) in order to control alternately the fixed target cycle and the four lepton cycles. All rectifiers are used in both cases and while this is not absolutely necessary for the lepton cycles it is more simple and requires less $du/dt$ per power supply. The two current loops are independent (separate references, DCCTs, etc.), so that they can be optimized separately. The only weak point of this solution is the quick changeover, which sometimes provokes a large jump in the voltage, but without negative consequences. A smooth changeover would certainly be preferable and will surely be developed soon.

For the collider operation, another loop is used, adapted to the coast level. The loop is optimized to minimize the final ripple, but nothing fundamental is changed.

4.5 Nuclear Magnetic Measurements

Measuring a homogeneous and constant field with an NMR is a well known technique. Several NMR probes are placed in one of the reference dipole magnets. Moreover, in our case, a special probe was built in order to measure the low field (0.01 T to 0.03 T). They allow accurate measurement ($\pm 10^{-5}$) of the local field. The whole range of the probes cover fields from 0.01 to 2.1 T. The part (a) of Fig. 18 shows the normal NMR signal as it is generated by a frequency modulation, while the probe measures a constant field.

The new idea used at the SPS consists of measuring the field during its rise by pre-programming fixed frequencies at a given time (for instance every 100 ms) and adjusting them accurately by using a "learning" program. This method, shown in part (b) of Fig. 18, allows complete scanning of a pulsed cycle with any time resolution. Such a measurement takes a few magnet cycles in order to be precise. Then, if required, a new measurement can be made by displacing the time origin, and so on. This solution is useful especially for the lepton cycles where, at 3.5 GeV/c, the magnetisation is difficult to determine due to the remanent field being about 10% of the required field.
5. OPERATIONAL ASPECTS

5.1 Interlocks

The repartition of the power supplies around the SPS accelerator could imply considerable risks for the personnel and materiel. It could have been arranged to switch off all 18 kV circuit breakers every time a fault occurred, but they are not able to withstand such a high number of operations between maintenance periods. Therefore, the following four-level interlock was chosen which aims to switch off circuit breakers as seldom as possible:

- **Level 1**: Protection of personnel
- **Level 2**: Protection of material (instantaneous damage)
- **Level 3**: Thermal protection (overload)
- **Level 4**: Warning

These different levels trigger the following actions:

- **Level 1**: Immediate 18 kV switch off of the faulty power supply.
- Inversion of all other power supplies, which fire the by-passes, in order to absorb the energy of the magnets quickly.
- Delayed switch off (2 s) of all the other circuit-breakers.

- **Level 2**: Immediate 18 kV switch off of the faulty power supply.
- Inversion of all other power supplies (with by-passes firing)

- **Level 3**: Inversion of all power supplies (with by-passes firing)

- **Level 4**: Simple warning to the control room.

Confidence in the correct operation of the 18 kV circuit breakers is limited. Therefore, the interlocks check if, after receiving an off order, the command has been obeyed. If not, the busbar circuit breakers are switched off after a delay of 2 seconds.

All faults are displayed on the front panel of the individual power supply (with the relevant level) and the information is transmitted to the control room. Several fault warnings are redundant (flow and temperature, for instance), so that it is possible to suppress one of them, if necessary, by putting a visible bridge near the fault lamp. Hidden "straps" are therefore definitively discouraged.

All the interlocks are powered by a small battery, so that it is always possible to read the interlocks locally for one hour after a complete blackout. Our experience has shown that it is very useful to preserve some local information since communication with the control room is often perturbed for some time after a blackout.

Every time a fault occurs, the fault status is recorded immediately and again 3 s later to see the consequences. No other post mortem system is installed at the moment, but it could be useful to have one. Each time the order to go into inversion is given to the rectifiers, the firing of the by-passes is checked and would prevent any new restart if all is not correct.
After switching on the power supplies, or after any fault, the firing pulses of all rectifiers are pushed to the invert limit. Before disabling this function, the position of all circuit breakers and all mechanical bypasses are compared with the configuration required by the program. If everything is not in order, all supplies are disabled.

The interlocks allow personnel to work on one faulty power supply while any of the others are running. This is achieved by isolating and grounding locally and completely the power part at the 18 kV input side and at the end of the passive filter by means of dedicated ground rods equipped with interlock contacts.

The following interlocks cannot be reset remotely, in order to enforce that a local check is first made:

- 18 kV fault (overcurrent, overload, homopolar)
- Emergency stop
- Overcurrent at the secondaries of a transformer
- Buchholz of a transformer
- Spark gap

5.2 Fault finding strategy

Operations personnel must understand the whole system though they do not need to be highly specialized in all fields. Some of the faults are easy to find and clearly presented. In this case, the decision to replace the deficient power supply by a spare can be made very quickly. The complete exchange can be made remotely from the control room. By eliminating a power supply, all the faults from interlock levels 2 to 4 are de-activated, although they can still be read. To avoid any risk of forgetting that a power supply is not ready to run, a "secret switch" must be kept on until all the repair work is completed and checked.

In some cases, intermittent problems occur. They are obviously much more difficult to solve. There are no automatic methods but the following routine can be followed systematically:

- Check the analog signals delivered by the power supplies to a central point near the control room. Eight signals are sent here from each power supply (voltages, reference, intermediate control signals) and facilitate a first diagnosis and determine whether there is a problem with one specific power or all of them.

- If there is a problem with one power supply, then replace it by another one and check if it functions correctly. If it does, the problem is solved but not understood, and the repair will be made later by specialists.

- If no fault can be observed on one specific power supply, then look at the common parts: reference, DCCTs, current regulator, timing, mains stability, load, etc. Check carefully whether there is any moment in the cycle where the fault repeats itself. If so, it could point to a relationship between the dc-current or dc-voltage amplitude.

- Then it starts to take time. A good approach consists of loading a cycle with about half of the maximum current with only half of the power supplies and then with the other half. If there is a difference, it could mean that one power supply is faulty. By making further mixings of six sets of equipment, the faulty one can be found. If no difference between one-half of the power supply and the
other could be found, then a comparison between the active and spare DCCTs could be made and common electronics exchanged. After this only the specialist can probably help, or someone with a very good intuition. Perhaps also a high voltage test of the magnets could show something abnormal.

- In the case of an earth fault, a central command makes it possible to localize the sextant of the tunnel in which it has occurred by using the by-pass switches. Switches have also been added to the quadrupole connections for this purpose.

- If a short circuit occurs in a magnet, it is not possible to detect it initially from the power supply but only from the beam behaviour. When it is approximately localized, measurement of the voltage across the magnet while pulsing and without beam allows the faulty magnet to be identified.

5.3 Difficulties

Some of the most relevant difficulties encountered during the 14 years of operation will now be described.

5.3.1 Mains problems on the 400 kV grid

Voltage drops can occur on one, two or three phases and all amplitudes of drop are possible. The most annoying ones are produced when thunderstorms are within a radius of 100 km from the site, the lightning often striking the high voltage line. Figures 20 and 21 show oscillograms of such perturbations. If it happens to occur during the descent (inverter mode), it is very difficult to ensure all commutations of the thyristors, and if only one is jumped, the destruction of one or more thyristors is highly probable. Many thyristors have been damaged under these circumstances. The presence of the compensators makes the task more difficult, since they can produce many unwanted harmonics at this moment. Sensitive and rapid voltage detectors were built which trigger the inverter protection and the by-passes as quickly as possible. This helps but cannot provide complete protection. During fixed target runs, the detection level is adjusted to 12%, and this is increased to 30% or even suppressed during the collider mode. Since the power available from the grid regularly increases, these problems decrease year by year. In 1989 for instance, not one thyristor was damaged because of mains voltage problems.

5.3.2 Thyristors

Numerous thyristors were damaged during the first few months of SPS operation for two main reasons. At the moment of switching on the 18 kV, it sometimes occurred that one thyristor self-fired due to the high rate of rise of the anode-cathode voltage. The thyristor still seemed to be in working order despite one spot of the silicon slice being damaged. However, after a while, the thyristor could not be fired anymore over its whole surface and it then short-circuited. This damage in two steps was difficult to find and required a great deal of investigation. The problem was cured by adding suppressor circuits. Although some thyristors installed at the beginning of the SPS are still running, this problem was also completely solved by natural selection of the thyristors and does not exist with more recent thyristors. There is no visible ageing of the thyristors. To the contrary, it seems that, after millions of pulses, the equivalent thermal resistance decreases.
5.3.3 Magnet connections

The dc connections of the magnets in the tunnel also created a serious problem for many years. As said before, they were not isolated but just held by insulated supports a few centimeters above the floor and protected by steel sheets. To avoid humidity, a small gap was left between the floor and this protection with the result that dust and small metallic pieces could very easily be pushed under the bars. After any shut-down, such metallic debris would be attracted by the pulsing currents in the busbars, finally creating a short circuit. Depending on the location of the short circuits, the voltage between bars can vary from zero to about 4 kV. Therefore, the consequences were very different varying from a sudden small anomaly in the current to a big flashover. In the first case it was difficult to find, in the second one long to repair. Detection of the short circuits was not possible using a simple earth detector since no earth current was flowing at the first moment; this only happened after the flashover occurred! A good detection method is to continuously measure the difference of the currents in the two loops of the dipoles, which is done with a magnetic device at the return point of the two loops, as shown before in Fig. 6. While it is possible to detect this fault very quickly, its location is much more difficult to find.

5.3.4 Rectifier transformers

The rectifier transformers caused trouble due to very small sparks occurring between the sheet copper secondary windings and the earth screen placed between the primary and the secondary, although the insulation level was theoretically high enough. While this problem could be solved by putting the screen at the same potential as the secondary, the screen caused more trouble than it was worth. The transformers are completely sealed and built without air dryers. If the machine is stopped for a few hours, the pressure inside the transformer becomes sub-atmospheric and air can be sucked in. Since the seals do not work very well under alternating pressure conditions, this leads to problems caused by the entry of air and humidity.

6. CONCLUSIONS

The basic technical choices have been confirmed to be correct. The SPS power supplies system works very efficiently and does not cause undue stoppages of the accelerator. However, a careful maintenance operation must still be completed once a year. After any long shut down of the machine it is very important to check all the power supplies very carefully, as is the habit, for 2 to 3 days before setting up the beam again. For the future, improvements in some of the hardware (circuit breakers, transformer seals, etc.) and the software support for fault diagnosis, including a clever post mortem memory, would surely help to save time in locating difficult faults.

ACKNOWLEDGEMENTS

Special thanks are due to all who have worked so hard to bring the main power supplies up to such good operating conditions. These include, O. Berrig, P. Campiche, H.K. Kuhn, M. Leroux and S. Oliver but above all S. van der Meer who was the initiator of the whole SPS power supply system.
Fig. 1 Fixed target cycle (dipoles)

Fig. 2 Lepton cycle (dipoles)

Fig. 3 Interleaved cycles (dipoles)
Fig. 6 Overall connections of the dipole magnets and power supplies

Fig. 7 Voltage distribution in a normal case

Fig. 8 Voltage distribution in emergency case
Fig. 9 Overall connections of the quadrupole magnets and power supplies

Fig. 10 One reactive compensator and harmonics filter

<table>
<thead>
<tr>
<th>Harmonic</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>5</th>
<th>7</th>
<th>11</th>
<th>13</th>
<th>17</th>
<th>17</th>
</tr>
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<tbody>
<tr>
<td>Q MVar</td>
<td>11</td>
<td>6</td>
<td>11</td>
<td>9</td>
<td>4.5</td>
<td>10</td>
<td>7.3</td>
<td>16.3</td>
<td>16.3</td>
</tr>
</tbody>
</table>

Fig. 11 Field and voltage control
Fig. 12 Field measurement

Fig. 13 B-Preset

Fig. 14 Typical bit error in the field measurement
Fig. 15 Diagram of the current and voltage control

Fig. 16 Diagram of the current control

Fig. 17 Diagram of the current control with two current loops

Fig. 18 Nuclear magnetic measurement
Fig. 19 Front view of an interlocks panel

Fig. 20 a) and b): Perturbation of the 400 kV and 18 kV networks during thunderstorms
CHOICE OF POWER SEMICONDUCTORS

J.-M. Peter
SGS Thomson

1. INTRODUCTION

We will consider in this paper that the power component operates like a switch (blocked, or conducting with a small voltage drop). In order to compare the components, we have to analyze:

- The possibilities in current and voltage (switchable power), permanent current, short-term overcurrent
- The switching behaviour concerning speed and losses
- The drive
- The cost.

The main components currently available (1990) are the power bipolar transistor (BJT), the power MOSFET, the insulated gate transistor (IGBT), the thyristor (SCR), the gate turn-off thyristor (GTO) and the power diode.

2. POWER SEMICONDUCTOR CHARACTERISTICS

2.1 Diode

The switchable power is defined by (see Fig. 1):
- Voltage limit (smaller than avalanche limit)
- Rated current.

The physical parameter limiting current is the maximum junction temperature.

*The maximum current in a diode depends essentially on the cooling: i.e. in practice on the thermal resistance (DC operating) and on the thermal impedance (short surge current).*

2.2 Bipolar transistor

The current possibility is defined by following these relations (see Figs. 2 and 3):

\[ V_{CE_{sat}} < 1.5 \quad \text{at} \quad I_C = I_{C_{sat}} \]
\[ I_C = I_{B_{sat}} \]

If \( I_C < I_{C_{sat}} \), the voltage drop \( V_{CE} \) follows a linear relation, and could be very low. However, if \( I_C > I_{C_{sat}} \), the transistor can be considered as a current source.

*The bipolar transistor has no overcurrent possibility (\( I_C > I_{C_{sat}} \)).

*The maximum operating current (\( I_{C_{sat}} \)) is defined by the gain, not by thermal considerations.*

*Voltage: Two parameters define bipolar transistor voltage capability:*

- \( V_{CE_{max}} \) max voltage with the base emitter junction blocked (\( V_{CE_{max}} = V_{CBO_{max}} \) max collector-base voltage)
- \( V_{CBO_{max}} \) max voltage with base open.

For switching applications voltage limits are defined by the Safe Operating Area (SOA) and the values for currently available components are shown in Table 1. (\( V_{CEW_{working}} \) working voltage at high current is often equal to \( V_{CBO} \)).
Table 1
Maximum bipolar transistor characteristics (1990)

<table>
<thead>
<tr>
<th>Type</th>
<th>( V_{CEO} ) [Volt]</th>
<th>( V_{CEV} ) [Volt]</th>
<th>( I_{C_{Sat}} ) [Amp]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Fast transistors</td>
<td>800</td>
<td>1300</td>
<td>60</td>
</tr>
<tr>
<td>Slow transistors</td>
<td>1000</td>
<td>1400</td>
<td>400</td>
</tr>
</tbody>
</table>

Voltage drop: If \( I_C < I_{C_{Sat}} \) the voltage drop (with adapted drive) is very low.
\[ V_{CE} = I_C/I_{C_{Sat}} \times V_{CE_{Sat}} \]

Drive (Fig. 3): During the conducting phase it is necessary to deliver a base current \( I_{B1} \) where
\[ I_{B1} = I_C/\beta \quad \beta = \text{gain} \]
At nominal current the specified gain is 10 for low-voltage transistors \( (V_{CEO} < 250 \, \text{V}) \) and 5 for high-voltage transistors. Near \( I_{C_{Sat}} \) (defined in the data sheets) the following empirical relation can be used to define gain at other current levels:
\[ \beta @ I_C = \beta @ I_{C_{Sat}} \times I_C/I_{C_{Sat}} \cdot \]

Switching time: It is necessary for a fast turn-off behaviour to force a negative \( I_{B2} \) current in the base. Two switch-off times are defined:
- Storage time \( t_s \). This is a 'memory' effect due to the minority carriers (1 \( \mu \)s and 3 \( \mu \)s for \( V_{CEO} = 100 \, \text{V} \) and 400 \( \text{V} \) respectively).
- Fall time \( t_f \). The majority of switching losses are due to the fall time
- Modern fast transistors (cellular technology) have very short fall times (Figs. 4 and 5).

2.3 Darlington

The major advantage of the Darlington (Fig. 6) is its higher gain but this advantage is counterbalanced by a higher voltage drop:
\[ V_{CE_{Sat \, \text{Darlington}}} = 0.8 \, \text{V} + V_{CE_{Sat \, \text{Base}}} \]
Another advantage is the possibility to operate it at a higher current density, because the gain of the power stage can be very low.

2.4 Thyristor

The thyristor (Figs. 7 and 8) is a positive feedback component. Consequently it has no saturation effect like the transistor.
\textit{The max operating current is defined (like a diode) by the cooling capacity}
\textit{The thyristor has very high current surge capabilities (within the thermal cooling time constant).}

Voltage: The thyristor blocking voltage can be very high (5 kV) as can be the rated current (2 kA).

Voltage drop: 0.8 V at very low current, 1.2 V at nominal current.
Drive: For turn-on (firing) the thyristor needs only a very low current during a short time, but the thyristor cannot be controlled for turn-off.

Switching times: When the anode current is forced to zero, the thyristor turns-off. But it is necessary to wait a time $t_{tr}$ (turn-off time) before reapplying the positive anode voltage.

2.5 GTO

The GTO (Fig. 9) is also a positive feedback component similar to the thyristor, but with an interdigital structure. Consequently it has a similar characteristic to the thyristor, but it can be blocked like a transistor.

Voltage: In 1990, some GTO's can support 4 kV with a maximum rated current of 1 kA. During the turn-off behaviour the maximum voltage is defined by the SOA which is poor at high operating currents.

Voltage drop: Similar to the thyristor but higher.

Drive (Fig. 10): For turn-off the GTO requires a very high negative gate current (gain = 3) leading to a sophisticated and expensive gate drive.

Switching times (Fig. 10): Like the bipolar transistor the GTO has a storage time, and during the fall time a tail effect increases considerably the turn-off losses.

2.6 MOSFET

This component (see Fig. 11) uses only minority carriers, which explains its specific behaviour. The majority carriers flow into the component by the influence of gate voltage, then current cannot be limited by a gain phenomenon. The voltage drop depends only on the silicon resistance $R_{ON}$.

The max operating current is defined (like a diode) by the cooling capability. The MOSFET has current surge possibilities (within the thermal cooling time constant) (Fig. 12)

Voltage: Due to the fact that the $R_{ON}$ increases considerably with the maximum rated voltage, the latter is presently limited to 1000 V. The MOSFET has a large SOA (it is able to sustain its maximum rated voltage during turn-off). Present performance values are:

<table>
<thead>
<tr>
<th>$R_{ON}$ at 25° (Ohms)</th>
<th>Max rated voltage (V)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.01</td>
<td>100</td>
</tr>
<tr>
<td>0.1</td>
<td>500</td>
</tr>
<tr>
<td>0.7</td>
<td>1000</td>
</tr>
</tbody>
</table>

Voltage drop: It is often claimed that the MOSFET has a very high voltage drop. This is not correct. The MOSFET voltage drop $R_{ON}I$ can be very low at low current density, but this is obtained at the expense of a large silicon surface area. The $R_{ON}$ is (unfortunately) specified in the international datasheets at 25°.

$$R_{ON} 100° = 1.7 \times R_{ON} 25°$$
Drive (Fig. 12): During conduction the gate requires only a voltage (= 15 V) without any consumption. MOSFETs turn-off very quickly (Fig. 13) when the gate source voltage reaches zero. If gate consumption during the conduction phase is zero, the designer has to consider losses due to charge (and discharge) of the gate source capacitance at each turn-on (turn-off).

Switching times: The MOSFET, majority carrier devices have no storage time. It is very important to remember this when considering applications of this device. Fall time (depending on the drive) can be very short. But for a rated voltage higher than 300 V, it is approximately the same as for fast bipolar and MOSFET transistors.

2.7 IGBT

The IGBT (Fig. 14) is the first child of the couple MOSFET-bipolar transistor. It can be considered as a pseudo Darlington with a MOSFET as the driver and a bipolar transistor as the power stage.

Maximum current is generally limited by cooling.
The IGBT has some overcurrent capability.

Voltage: The present maximum rated voltage is 1.2 kV. This limit will rise next year (1.8 kV) as well as the maximum rated current (500 A). The SOA is quasi-rectangular.

Voltage drop: $V_t = 1.3 + r_t$ for the same silicon surface (and the same rated voltage). The IGBT voltage drop at high current density is much smaller than that of the MOSFET (same silicon chip area). But, at low current density, the IGBT always has relatively high conduction losses which limit the efficiency of IGBT equipment.

Drive: Similar to the MOSFET drive.

Switching times (Figs. 15 and 16): The MOSFET has practically no storage time. But like the GTO, a tail current (due to uncontrolled turn-off behaviour of the bipolar part) increases switching losses.

3. FAST POWER DIODES

The majority of power semiconductors operate with fast power diodes (free wheel, rectifier, clamping, etc.). Power PN diodes have a memory effect due to minority carriers. When the diode current decreases and reaches 0, the diode is still full of minority carriers, and can be considered as a short circuit during a short time. The inverse current due to this effect is a source of trouble (peak currents, noise, overvoltages, supplementary losses). Knowledge of fast recovery diodes is therefore important in the design of power circuits. The essential features are:

- Turn-off behaviour as shown in Fig. 17. The main parameter is the inverse current $I_{RM}$, and in some cases it is the recovery charge $Q_r$.
- Inverse current increase with $dI/dt$ (slope of decreasing current before turn-off) and with $T_j$.
- A fast PN diode is one with a reduced life time of the minority carriers which leads to a reduction of the diffusion length. If the latter is shorter than the silicon thickness of the N region, the diode-on resistance increases drastically. The design of a fast diode is therefore the result of a trade-off between maximum voltage $V_{DRM}$, voltage drop $V_F$ and speed ($I_{RM}$). Figures 18 and 19 show the state of the art for several 12 A fast diodes.
The Schottky power diodes using only majority carriers have a different behaviour, smaller voltage drop and no recovery charge. These advantages are obtained at the expense of:

- Reduction of maximum voltage capability (60 ... 100 V)
- High internal capacitance.

Figures 20 and 21 show losses introduced by a free-wheel diode. Using a faster diode reduces these losses, but it is not always possible to have an ultra-fast diode with high voltage. A useful method consists in using several low voltage ultra-fast diodes in series (see Bibliography). When the diode switches-off in series with an inductance L a supplementary energy $1/2L I^2_{RM}$ is dissipated in the circuit (Fig. 22). For these reasons the choice of the circuit configuration is important.

4. **LIMITS AND MAXIMUM RATINGS**

   The absolute maximum ratings are defined by the semiconductor manufacturer and are illustrated in Fig. 23. If these are exceeded the component could be destroyed. Absolute maximum ratings are applied to $T_j$, torque, voltage, SOA, $I_{RM}$ and $I_{max}$.

   **Attention:** The user cannot measure an absolute maximum rating since there is always a chance that the component will be destroyed. However, the other characteristics such as $V_{CEsat}$, weight, dimensions, switching times ... can be measured. It is important to take into account the worst case (maximum or minimum value) and to verify that the circuit operates well within the spread of all the parameters.

5. **COST OF THE SWITCH FUNCTION**

   In order to do a comparison, the designer has to analyse not only the price of the component, but also the function cost:

   
   \[
   \text{Component price} + \text{Drive (+ protection) cost} = \text{Total function cost.}
   \]

   The manufacturing cost of the semiconductor power component depends on:

   - the complexity of the process
   - the silicon surface
   - the quantity manufactured.

   The indication given in Fig. 24 could be very useful to define the long-term trends.

6. **COMPARISON OF POWER SEMICONDUCTORS**

   Table 2 summarizes the essential characteristics of the power semiconductors and we can deduce the following:

   1) The power MOSFET is almost the perfect component, it has all the advantages (apart from the parasitic diode and the high DS capacitance). However, its cost could be very high since it needs a high-silicon surface, especially if $V_{DS} > 400$ V.

   2) The main advantage of the bipolar transistor is the cost ($V > 400$ V) but there are the disadvantages of storage time and the necessity to develop an adapted base drive. In the majority of applications the cost of the drive is higher than the cost of the component.
3) The IGBT has the advantages and disadvantages of the 'parent' (MOSFET-Bipolar) and is in competition with the bipolar transistor. There are two major advantages:
- no storage time
- low cost of the drive
and two disadvantages:
- tail current at turn-off (switch-off losses)
- high voltage drop at low current density.

4) The GTO is adapted to high power only (I > 200 A) especially if the maximum voltage is higher than 1.2 kV.

5) The thyristor (triac) is the cheapest component; when the phase control drive is adapted it is impossible to find a more economical solution.

<table>
<thead>
<tr>
<th>Device</th>
<th>Advantages</th>
<th>Disadvantages</th>
</tr>
</thead>
<tbody>
<tr>
<td>Power bipolar</td>
<td>Component cost V &gt; 300 V, very important V &gt; 700 V</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Resistive behaviour (I &lt; I_{lim})</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Short fall time (new technology)</td>
<td>Cost of the drive</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Storage time</td>
</tr>
<tr>
<td></td>
<td></td>
<td>No overcurrent possibility</td>
</tr>
<tr>
<td></td>
<td></td>
<td>R BSOA limited to V_{CEW}</td>
</tr>
<tr>
<td>Power MOSFET</td>
<td>Low cost of the drive</td>
<td>Component cost V &gt; 300 V</td>
</tr>
<tr>
<td></td>
<td>Overcurrent possible</td>
<td>Component high cost (V &gt; 700 V)</td>
</tr>
<tr>
<td></td>
<td>No storage time</td>
<td>Parasitic antiparallel diode</td>
</tr>
<tr>
<td></td>
<td>Resistive behaviour</td>
<td>DS capacitance</td>
</tr>
<tr>
<td></td>
<td>Very short switching times</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Rectangular R BSOA</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Some case: antiparallel diode</td>
<td></td>
</tr>
<tr>
<td>GTO</td>
<td>Component cost (high power)</td>
<td>Cost of the drive</td>
</tr>
<tr>
<td></td>
<td>Turn-off control</td>
<td>Storage time</td>
</tr>
<tr>
<td></td>
<td>Over current possible (moderate)</td>
<td>No overcurrent possibility</td>
</tr>
<tr>
<td></td>
<td></td>
<td>R BSOA limited to V_{CEW}</td>
</tr>
<tr>
<td>IGBT</td>
<td>Low cost of the drive</td>
<td>Component cost V &gt; 300 V</td>
</tr>
<tr>
<td></td>
<td>Component cost</td>
<td>Component high cost (V &gt; 700 V)</td>
</tr>
<tr>
<td></td>
<td>Over current possible (moderate)</td>
<td>Parasitic antiparallel diode</td>
</tr>
<tr>
<td></td>
<td>No parasitic diode</td>
<td>DS capacitance</td>
</tr>
<tr>
<td></td>
<td>Quasi rectangular R BSOA</td>
<td></td>
</tr>
<tr>
<td></td>
<td>No storage time</td>
<td></td>
</tr>
<tr>
<td>Thyristor</td>
<td>Component cost: very low</td>
<td>Tail effect ... &gt; off losses!</td>
</tr>
<tr>
<td></td>
<td>Low cost of the drive</td>
<td>Always voltage drop</td>
</tr>
<tr>
<td></td>
<td>Overcurrent possibilities</td>
<td></td>
</tr>
<tr>
<td></td>
<td></td>
<td>No turn-off control</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Limited t_q (8-20 μs)</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Voltage drop</td>
</tr>
</tbody>
</table>
7. CONCLUSION

"Power MOSFET has very high voltage drop"
"Bipolar ... an old technology"
"Epitaxial is better ..."

Such commercial advertising is not able to help a designer to optimize equipment. There is a very wide choice of components and no 'best' solutions, we are in the field of technology and not in the field of science. If for some applications, for instance 1 MHz SMPS, only one solution (MOSFET) is possible, for most applications there are always several solutions; the designer's job is to optimize the switch function after a thorough analysis. Our experience shows that the quality of this analysis, and the work done by the designer (drive, protection, etc. ...) has a greater importance on the total cost than the price of the component itself.

* * *

BIBLIOGRAPHY

Transistors and Diodes in Power Processing, Thomson Semiconductor 1985, Chaps. 1, 2, 3, 4, 6, 7, 8, 9, 11, 15, 19, 20

Fig. 1 Power diode

\[ V_F = \eta + pI \]
\[ P = eI_{AVG} + pI^2 \]

Fig. 2 Bipolar transistor

\[ V_{CE sat} < 1.5V \]
\[ I_C = I_{C sat} \]
\[ I_B = I_{B sat} \]

Fig. 3 Bipolar transistor: how to drive

Gain

How to choose base drive current

Fig. 4 Switch-off behaviour of two high current transistors from 50 A (T1 = 45 °C).

In each case optimal base drive.
A modern high voltage transistor (V_{CE} 450 V, V_{CE} 850 or 1000 V) realised with an 'ETD' (Easy to drive) technology has (at 100 °C) a fall time less than 50 ns.

Fig. 5 New cellular structure 'ETD transistor reduces switching losses'

The Darlington (monolithic or hybrid) gives for a same chip surface more current than the equivalent transistor.

The second stage is operating with lower forced gain at higher current density. In the given example T2 is a transistor normally specified at 10 A with a gain of 3.

Near I_{Dest} the product n_{p} A_{p} = constant.

Fig. 6 Bipolar Darlington

<table>
<thead>
<tr>
<th>Current limit</th>
<th>Voltage limit</th>
</tr>
</thead>
<tbody>
<tr>
<td>I_A \approx \frac{k}{1 - \beta_1 \beta_2}</td>
<td>V_T = V_o + P I</td>
</tr>
<tr>
<td>P = \frac{e_o T_{AVE} + P_{T}}{T_{RMS}}</td>
<td>V_{Thermal} = V_{Source}</td>
</tr>
<tr>
<td></td>
<td>Switching SOA</td>
</tr>
</tbody>
</table>

Turn-off:
1) Force the anode current to reach 0 by external circuit
2) Apply a negative anode-cathode voltage during a time ≥ t_{on}

Fig. 7 Thyristor

Fig. 8 Thyristor: How to drive
<table>
<thead>
<tr>
<th><strong>Current limit</strong></th>
<th><strong>Voltage limit</strong></th>
</tr>
</thead>
<tbody>
<tr>
<td>$I_A = \frac{K}{1 - \beta_1 \beta_2}$</td>
<td>$V_T = e_0 + p I$</td>
</tr>
<tr>
<td>Positive feedback</td>
<td>l) Thermal limit</td>
</tr>
<tr>
<td>$\beta_1 \beta_2 &gt; 1 \rightarrow$ on</td>
<td>2) $I &lt; I_{C-L}$</td>
</tr>
<tr>
<td>$\beta_1 \beta_2 &lt; 1$ by gate control</td>
<td>Switch aid network necessary</td>
</tr>
</tbody>
</table>

**Fig. 9** GTO

**Turn-off gain:**

$$\frac{I_A}{I_{6\theta}} \approx 3$$

**Fig. 10** GTO: how to drive

<table>
<thead>
<tr>
<th><strong>Current limit</strong></th>
<th><strong>Voltage limit</strong></th>
</tr>
</thead>
<tbody>
<tr>
<td>$I_D$</td>
<td>$V_G$</td>
</tr>
<tr>
<td>$V_{DS} = R_{on} \cdot I$</td>
<td>$I_{G\delta}$</td>
</tr>
<tr>
<td>$P = R_{on} \cdot I_{RMS}^2$</td>
<td>$I_{G\delta}$</td>
</tr>
<tr>
<td>Attention!</td>
<td>$V_{DS}$</td>
</tr>
<tr>
<td>$R_{on}(125) = 1.8 R_{on}$</td>
<td>$V_{GSS}$</td>
</tr>
</tbody>
</table>

**Fig. 11** MOSFET transistor

**Don't forget the gate source capacitance.**

*Gate voltage: +15V
*Logic level Mosfet: +5V

**No storage time**

**Fig. 12** MOSFET: how to drive
Fig. 13 Turn-off behaviour with high current

Nominal current

Very high surge current

Fig. 14 IGBT

Higher voltage drop than
- Darlington
- Thyristor

Fig. 15 IGBT switch-off behaviour

Fig. 16 IGBT: how to drive
TURN-OFF BEHAVIOR

Different diodes: 200 V, 400 V, 800 V

CONSEQUENCES: Current spikes, noise, supplementary losses

Fig. 17 Diode turn-off behaviour

Fig. 18 Different diodes: 200 V, 400 V, 800 V

Fig. 19 Switch-off behaviour versus maximum voltage rating

The design of a fast diode is the result of a compromise between maximum forward voltage drop and reverse current. Each diode has a 12 A rated current, and the inverse current is measured at 100 V in the same operating conditions.
The diode D turns off when the transistor turns on.
With a medium fast diode BYX66, the transistor turn-on losses are very high.
If one uses faster diodes, the losses are very small. One can notice that the diodes' losses are always small in comparison with the transistor turn-on losses.

Fig. 21 Free-wheel diode turn-off.

Fig. 22 Two different behaviours. Why?

Fig. 23 Maximum ratings and characteristics

Fig. 24 Chip surface versus maximum rated voltage
WOUND COMPONENTS

V. Richter
Starkstrom-Gerätebau GmbH, Regensburg, FRG

ABSTRACT
Transformers and reactors, collectively referred to here as wound components, have gained a high reputation in efficiency, manufacture, quality control and testing. By means of computer programs it is possible to calculate many technical parameters with the result that values obtained in tests are in very good agreement with the calculated figures. Optimum technical solutions can result in economies by way of automation in production. Serious efforts have been made over the last years to improve materials in order to reduce core losses and so improve efficiency. In addition to traditional liquid-immersed transformers and reactors, dry types up to 15000 kVA and Um 36 kV are available in cast-resin design and have excellent properties.

1. INTRODUCTION

In this paper some of the technical problems and their solutions in connection with the design and manufacture of transformers and reactors will be described for a range starting from, say 100 kVA to some MVA. The following subjects will be treated in detail:

- cores (types, materials and manufacture)
- windings (types, conductor construction and materials)
- cooling (dry, liquid-immersed)
- special problems and requirements of converter transformers.

The above mentioned kVA range corresponds to the power needed for ac-to-dc converters for accelerators. The special requirements for transformers and reactors in this field are different to those in the usual distribution systems. Criteria which should be considered are:

- pulsed loads
- frequent switching
- high dynamic stress
- frequencies higher than 50 or 60 Hz
- high currents
- harmonics in currents and voltages
- superposition of a.c. and d.c. components in currents and voltages
- sophisticated winding connections and arrangements
- multi-winding designs
- screens between HV and LV windings
- requirements of symmetry of voltage, resistance and reactance between phases and windings
- consequences of failures in power converters
- special accessories.
These criteria can create many problems for designers and users of such equipment. However it will be pointed out that non-standard components are very well able to comply with the special requirements. But, generally speaking, transformers and reactors for power converters are specially designed for a particular project and are only produced as a small series. Highly qualified people are needed in the design departments, for the manufacturing process and for quality control. Taking into account these facts, it is easy to understand that the cost must be much higher than the standard products used in the usual distribution systems. On the other hand, these products work in extremely complicated and expensive systems, so that quality and safety aspects must be brought into the foreground.

2. **CORES**

2.1 **Materials**

During the gradual development of materials for cores, there were times when great progress in the reduction of losses, i.e. hysteresis loss and eddy current loss was achieved. Although no-load losses in transformers and reactors in distribution systems are of much greater importance, this development is also important for special transformers.

In the early stages of transformer production, non-oriented core material was used, as is still the case today for rotating machines, reactor cores and magnets. Since in the case of power transformers the amount of core loss is equal to the no-load loss, it is important to keep the core losses low; the eddy currents are reduced by using the core material in the form of thin sheets of a thickness of 0.35 mm.

The first big step in loss reduction in cores was made by introducing cold-rolled, grain-oriented, steel; the production method was invented in 1934 by Goss [1] in U.S.A. If the magnetic flux direction is parallel to the rolling direction, superior magnetic properties, i.e. low loss and magnetizing power, can be achieved. The material was produced with a thickness of 0.35 mm, nowadays it is available in thicknesses of 0.3 and 0.27. This material is referred to as conventional grain-oriented (c.g.o.) core steel in order to distinguish it from the next step in development, which started in the late 1960's with the introduction of the HI-B qualities (high permeability) in Japan [2].

At this point, the different factors by which the two loss components of grain-oriented material are affected, should be mentioned:

i) Hysteresis loss by:
   - grade of grain orientation
   - purity of bulk material (inclusions, metallurgy)
   - internal mechanical stress

ii) Eddy current loss by:
   - electrical resistivity (Si content)
   - size of magnetic domains (grain size)
   - surface tensile stress (coating)
   - thickness.
The improvements were achieved by increasing the degree of orientation, higher Si content and special surface treatments to introduce tensile stress. The main results were lower excitation power, lower loss especially at flux densities over 1.5 T, and lower magnetostriction values. The effect of magnetostriction has a large influence on the noise generation of transformers. In addition, it should be mentioned that mechanical stress introduced during the cutting process and core manufacture, has a much lower influence on the deterioration of magnetic properties in the case of the HI-B material which has now been produced in Europe for about 15 years and with a thickness of 0.3 mm.

New efforts made in Japan resulted in further improvement of the HI-B qualities by affecting orientation and surface condition, by increasing the Si content, by reducing the thickness to 0.23 mm, and by refining domain wall-spacing. The size of the domains affects the anomalous loss. Abrasion and laser (or plasma) irradiation of the surface of HI-B material for refining domain wall-spacing is used. These new HI-B qualities with a thickness of 0.23 mm have been available in Japan for about 5 years and in Europe for 1 year [3,4].

Development of a different type started in the U.S.A. in the late 1970's and resulted in the amorphous metals or metallic glasses which have since become well known. A molten FeB alloy is fed onto a rotating, cooled wheel, where it solidifies in approximately 1 milisecond to give a continuous, thin ribbon. As there is no crystalline structure and its thickness is about 0.025 mm, losses are only about 1/5 of that of the HI-B material. The width of ribbon is limited to about 200 mm and in its original thickness it is only possible to use this material for wound cores. Further main disadvantages of this material are brittleness, a saturation flux density of 1.6 T and a lamination factor of about 0.85 in comparison with 2.0 T and a value of 0.97 for grain oriented materials. In order to use this material in a similar manner to the oriented materials in stacked cores, a method was found to combine several of the thin ribbons into a 0.13 mm strip. Several small distribution transformers with ratings up to about 400 kVA have been built in Europe by different companies for assessment purposes [5,6].

A survey of loss and magnetizing power of the different materials and steps of development is given in Table 1. The figures given are for laboratory measurements carried out in accordance with IEC standards for example.

<table>
<thead>
<tr>
<th>Material</th>
<th>Thickness (mm)</th>
<th>Specific Losses (W/kg)</th>
<th>Specific Power (VA/kg)</th>
<th>Magnetizing Power (VA/kg)</th>
<th>Remarks</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cold-rolled non-oriented</td>
<td>0.35</td>
<td>0.95 2.35 3.1</td>
<td>1.8 48.0</td>
<td></td>
<td>Best quality</td>
</tr>
<tr>
<td>Conventional grain-oriented cold-rolled (e.g.o.)</td>
<td>0.35</td>
<td>0.42 0.98 1.42</td>
<td>0.49 1.25</td>
<td></td>
<td>Standard quality</td>
</tr>
<tr>
<td></td>
<td>0.3</td>
<td>0.38 0.89 1.42</td>
<td>0.49 1.25</td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>0.27</td>
<td>0.35 0.81 1.18</td>
<td>0.42 1.1</td>
<td></td>
<td></td>
</tr>
<tr>
<td>HI-B</td>
<td>0.3</td>
<td>0.39 0.88 1.21</td>
<td>0.45 1.05</td>
<td></td>
<td>Standard quality</td>
</tr>
<tr>
<td></td>
<td>0.34</td>
<td>0.34 0.76 1.05</td>
<td>0.45 1.05</td>
<td></td>
<td>Best quality</td>
</tr>
<tr>
<td></td>
<td>0.23</td>
<td>0.29 0.65 0.90</td>
<td>0.34 0.76</td>
<td></td>
<td>Domain refined</td>
</tr>
<tr>
<td></td>
<td>0.23</td>
<td>0.25 0.59 0.80</td>
<td>0.35 0.94</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Amorphous metals</td>
<td>0.025</td>
<td>0.07 0.13*</td>
<td>0.18 0.3*</td>
<td></td>
<td>Original ribbon strip (several plies)</td>
</tr>
<tr>
<td></td>
<td>0.15</td>
<td>0.10 0.18*</td>
<td>0.18 0.3*</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

*) at 1.3 T
2.2 Core cross section

Typical forms of design of core cross section are shown in Fig. 1. Rectangular cores as shown in Fig. 1a are used for small transformers and reactors. However, designs exist also in the range of some MVA. In this case the economics of core manufacture are in the dominating foreground. If it is possible to fit the windings very close to the rectangular core, a very economic solution with respect to space and weight of the active part can be found. However, the manufacture of rectangular windings and, much more important, their ability to withstand short-circuit conditions raise many problems.

Usually transformers are designed with cruciform cores, as in Fig. 1b, for a wide range of power. The effective iron cross section can be increased by increasing the number of different widths within a given diameter. The actual number used is a compromise between the effective cross section required and the economics of manufacture. Modern cutting machines stack complete legs and yokes automatically, so an increase in the number of widths is possible without a large influence on the costs.

Perfectly circular cores with laminations in the radial direction as in Fig. 1c are used today especially for reactor core packets. The effective core cross section is smaller than for cruciform cores. In the neighbourhood of air gaps, magnetic flux enters the core packets in the radial direction with the result that the loss, especially at higher frequencies, is much lower [7].

![Core Cross Sections](image)

Fig. 1 Typical cross sections

2.3 Core types

The design of cores used for transformers or reactors is influenced by the following factors:

- type of transformer or reactor
- rating
- acceptable core height
- manufacturing possibilities and economic considerations.

A complete core is formed by the main limbs with their surrounding windings, auxiliary limbs without windings, and yokes which connect the limbs. Usually a three-phase transformer or reactor will have three main
limbs while a single-phase design will have only two. A single-phase design may also have one main limb and one or two auxiliary limbs. In the latter case the cross section of the auxiliary limbs and yokes may be only half that of the main limb.

In the case of transformers, the yoke cross section can be the same as in the limbs, which is a great advantage for automatic production. For reactors, the core consists of several core packets and air gaps; there is no joint between the limb and yoke laminations. The cross section of the yokes can be different to that of the limbs, usually higher, rectangular designs will be used. Magnetic flux in the yokes is higher than that in the limbs since a certain amount of flux which enters the yokes comes from the space between the winding and core and in the winding itself. A reactor can also be designed without any magnetic core packets, i.e. with only yokes (air-core inductor). Figures 2a-f show some different core types for three-phase and single-phase application.

![Fig. 2 Core types](image)

**2.4 Joints**

Reactors of the usual design with core packets and air gaps have a butt joint with a very small air gap to facilitate axial compression (Fig. 3a). Pressure is necessary to avoid the vibrations caused by magnetic forces in the air gaps. The air gap distance must be maintained with a material having a high modulus of elasticity, for example porcelain.

For transformers, type and quality of the joints have a large influence on losses, magnetizing power and noise generation of the core. If non-oriented laminations are to be used, simple butt and overlap joints, according to Fig. 3b, are adequate. For oriented material with superior magnetic properties in the rolling direction, it is
important to minimise the regions with deviating flux directions. It was found that a 45 degree cut is the best solution for this purpose, Fig. 3c.

![Joint arrangements](image)

a) Butt joint (reactor)  
b) Butt and overlap joint (non-oriented material)  
c) 45 degree overlap joint (grain oriented material)

Fig. 3 Joint arrangements

The ratio between measured core loss of the completed transformer and the value calculated with the figures of Table 1 (laboratory measurements), is called the "building factor". The following main effects increase this factor:

- deterioration of material properties by mechanical stress during the cutting process
- deviation of flux direction from orientation of the material near joints
- compressive stress, introduced during core manufacture by core-clamping structures or by waviness of laminations
- increase of flux density in areas adjacent to the small gaps at joints
- eddy currents in the region of the overlap where flux crosses over to the next layer before and after the small gaps at the joints.

By means of a very special overlap technique, the so-called 'step-lap', it is possible to limit the increase of flux density in the joint areas, so reducing core loss, magnetizing current and noise. Figure 4 shows the principle of the step-lap in comparison with a standard overlap. It is obvious that this construction can only be achieved exactly and economically by means of a special machine which produces complete legs and yokes.

![Lamination overlap](image)

a) Standard overlap (two positions)  
b) step-lap (example with five positions)

Fig. 4 Lamination overlap
In comparison with standard overlap the best results with step-lap cores are achieved in a flux density range between 1.4 and 1.6 T. With a 630 kVA distribution transformer a reduction of 5% in no load losses, 50% in magnetizing power and 5 dB(A) in noise can be obtained.

At this point it is important to mention that improved magnetic properties increase remanent flux density by about 30% with the result that the peak in-rush currents become higher with step-lap cores.

3. **WINDINGS**

3.1 **Materials**

Copper and aluminium are the two materials which are used as conductors. Usually copper is preferred because of its higher conductivity and better mechanical properties. Aluminium has its place for special conductor types such as foils.

Materials for the inner insulation of coils and for winding support structures are very different in the case of oil or liquid-immersed types compared with dry types. As the average temperature rise of windings is limited to 65 K in the case of oil-immersed types, materials such as paper, pressboard, enamel and wood can be used.

The heat transfer between a winding and its cooling liquid is 10 to 15 times better than that between a winding and air by natural convection and, if possible, by thermal radiation. Therefore, the size of air cooling ducts has to be at least twice to three times larger than in oil-immersed types. To limit the increase of size for dry types it is obvious that the admissible temperature rise has to be increased significantly. Special materials such as foils, fibreglass, resin-impregnated fibreglass and enamels are available for an admissible temperature rise up to 125 K and more.

In all cases the temperature at a "hot spot" of a winding determines the life expectancy of a transformer or reactor. Materials have to be selected carefully and ageing tests of the complete insulation system performed to assign materials to a special temperature index (allowable temperature rise) especially for the dry types.

3.2 **Conductors**

The following conductor types are used:

- round wire (enamel coated, sometimes paper covered)
- rectangular conductors (paper, enamel or glass insulated)
- foils (without any particular conductor insulation)
- transposed cables (strands enamel coated, complete bunch paper or glass-ribbon covered)
- hollow conductors (rectangular, for direct cooling of conductor, for example by water).

3.3 **Winding types**

In modern transformer and reactor design and manufacture, special types of windings have been established in the kVA range covered in this paper. On the one hand, easy and economic manufacture should be achieved, if
possible in an automatic way, on the other, problems should not arise due to mechanical and electrical stress during tests or under continuous service conditions. Figure 5 shows the most important types of windings.

![Diagram of winding types]

a) layer winding  
b) sheet (foil) winding  
c) (continuous) disc winding

Fig. 5 Winding types

Layer windings, illustrated in Fig. 5a, are very often used for both dry and liquid-immersed types. Rectangular and round conductors as well as transposed cable can be used, depending on current and kVA size, while in special cases hollow conductors can also be used. High-voltage (HV) windings of distribution transformers, cast-resin transformers, rectifier transformers and reactors up to a maximum system voltage of 36 kV take the form of layer windings. Should lightning-impulse stress occur, layer windings ensure an optimal voltage distribution with few oscillations. It is important to mention that axial cooling ducts can be provided very easily for both dry and liquid-immersed types.

Foil or sheet windings as shown in Fig. 5b are typical for low-voltage (LV) windings of transformers up to 1 kV. Foil windings are well adapted to high currents and offer many advantages in withstanding short-circuits, effects of high frequencies or harmonics on eddy-current losses, while offering economic manufacture including the cooling ducts. However, some problems can arise if sheet windings are used for reactors with gapped cores. If the distance between the core and winding is small compared to the axial size of the gaps, the radial flux, which is apparent near the gaps, can produce high local eddy currents and losses, especially at higher frequencies. It should be mentioned that radial stray flux will always cause eddy currents in foil windings. On the other hand, the radial flux is reduced by these eddy currents, which means that foil or sheet windings act like an electromagnetic screen.

Disc windings, as shown in Fig. 5c, are used for system voltages higher than 36 kV and are here always of the liquid-immersed type. Rectangular conductors and transposed cables will be provided at these voltages. In the case of dry transformers this type of winding can also be used for HV-windings up to and including 36 kV, especially with small foils or round conductors embedded in cast resin. However, the cooling effect of horizontal and vertical (axial) cooling ducts will be reduced by the spacers necessary between the discs. For cast-resin designs it is very difficult to provide axial cooling ducts at all.
Disc windings can also be used for high currents and extremely low voltages by connecting two discs in series, and all double discs in parallel, to heavy busbars. Usually the LV winding is placed adjacent to the core, the HV winding outside. However, for high currents the disc winding has to be the outer one.

4. **COOLING SYSTEMS**

There are two main cooling systems for transformers and reactors:

- oil immersed, including mineral and synthetic insulating and cooling liquids such as PCB, synthetic esters and Silicone oil
- dry, where core and windings are not immersed in an insulating (and cooling) liquid but are cooled by air or gas.

Circulation of the cooling medium may be natural, or forced by pumps or fans. For liquid-immersed transformers a secondary cooling medium, usually air or water, must always be provided; the secondary cooling medium may also circulate naturally or by force.

For special purposes dry types can also be designed as sealed units (like oil-immersed types), in such a way that there is no exchange between the primary cooling medium (air or gas) and the secondary cooling medium (air or water); primary as well as secondary cooling fluid circulation may be natural or forced. Therefore, dry types can now fulfil nearly all the requirements which were previously only possible with oil-immersed types.

Designs with hollow conductors look like dry types, but the cooling is very efficient; the fluid is usually water. Compact windings without any cooling ducts and high current densities can be achieved; typical examples are reactors for high currents and/or high frequencies.

At this stage it would be helpful to point out the advantages and disadvantages of dry types, especially cast-resin designs.

**Advantages are:**
- Installation close to load centre, because of low fire risk
- no risk of the cooling liquids contaminating the environment
- no sumps or basins for cooling liquids, which means high flexibility for choice of location
- high short-time-overload capacity (because of low current densities)
- ability to withstand extremely high short-circuits

**Disadvantages are:**
- highest voltage for equipment limited to 36 kV
- rated power limited to about 15000 kVA
- outdoor installation only possible with special enclosure or as sealed type (for extreme ambient conditions)
- higher weight for active part (core and coils)
- higher cost.
During the last few years many old PCB-filled transformers have had to be replaced, in many cases by cast-resin dry-type transformers. We now find the modern dry type in industrial service, airports, hospitals, industrial zones as well as in urban supply systems and in special areas where water contamination has to be prevented. Therefore, substitutes for PCB such as Silicone oil or synthetic esters are of minor importance.

5. **RECTIFIER-TRANSFORMER DESIGN**

5.1 **Dynamic stress**

Due to high overloads, pulse loads or frequent short-circuits, rectifier transformers often are exposed to much higher dynamic stress than standard distribution transformers. Load currents generate stray magnetic flux within the windings and their surrounding structure. Thus, the conductors within the windings in a standard two-winding transformer are subjected to forces such, that axial flux density creates radial forces, and radial flux density axial forces. At the rated current these forces are usually small. But, since they increase in proportion to the square of the current, they have some effect at high overloads or with pulse loads and become extremely high for the first current peak of asymmetrical short-circuit currents.

According to their effect and direction, three different kinds of forces have to be considered:

- radial forces
- axial symmetrical forces
- unbalance forces

As the direction of the currents is opposite in the LV and HV windings, the winding adjacent to the core is subjected to compressive radial stress, the outer winding to tensile stress. To avoid permanent (plastic) deformations, the stress in the conductors has to be limited to match the properties of their material. Compressive stress also raises the problem of buckling which largely depends on the radial thickness of the conductor and the number of radial supports. This means that foil or sheet windings, often used as the inner winding in both oil-immersed and dry types, could be seriously affected. To solve this problem, a special resin pre-impregnated inter-turn insulation is used. After completion of the winding work, the resin is cured by heating with the result that all turns form a strong tube. The rigidity of the winding should be subject to quality control [8-10].

Axial symmetrical forces, caused by radial flux components, try to reduce the axial dimensions of both the LV and HV windings. This will result if the conductor insulation, spacers and other insulating structures in the windings change dimensions elastically or by permanent deformation. In time, continuous movements of the conductors or complete windings may result in failure of the transformer. It is not important whether this arises from high forces during short-circuits, or moderate forces during a longer time in the case of overloads or pulse loads. To prevent movements, windings have to be clamped according to the forces acting on them. In most designs core clamping structures will also be used to keep the windings under a certain axial pressure [11].

Instead of providing heavy clamping structures, it would seem better to use windings which cannot change their dimensions under the influence of axial compressive forces, e.g. foil windings. In this case, the conductor covers the whole axial dimension of the winding. Radial flux components are reduced by eddy currents in the sheet, as already mentioned in section 3.3. The compressive forces acting on a sheet winding are consequently
much smaller than for a layer or disc winding of the same dimensions under the same short-circuit conditions. Windings cast as a solid block give the same advantage; axial compressive forces have no effect on the axial winding dimensions.

Axial unbalance forces always occur if the distribution of turns between the HV and LV windings is not symmetrical (so-called ampere-turn unbalance). This can be caused, for example, by misalignment during assembly of the coils on the core, or by an asymmetrical arrangement of taps in HV windings. In either case additional radial flux components are produced. Moreover, the resulting forces try to increase the unbalance so that the forces increase and so on. This means that clamping structures have to be dimensioned to accept the expected unbalance forces. If foil windings are used, radial flux components are reduced and, as a result, also the unbalanced axial forces. This is explained by the fact that eddy currents, caused by radial flux, superimposed on load currents, form a more symmetrical ampere-turn distribution between the HV and LV windings. In other words, the current distribution in sheet windings is a mirror image of the distribution in the other winding [12,13]. Windings individually cast in resin cannot avoid unbalance forces. If both the HV and LV windings are cast in one block, the unbalance forces do not act on the clamping system.

Designing power transformers and especially rectifier transformers with LV sheet windings provides high safety margins with respect to dynamic performance under short-circuit conditions. If cast-resin transformers are selected for a power converter project, optimal performance can be expected as explained above.

5.2 Effect of harmonic currents

In the previous section forces produced by stray flux were described. Stray flux also causes additional losses in the form of:

- eddy currents in windings
- eddy currents in all kinds of structures made of conductive material (clamping frame, tank)
- eddy currents in screens and parts of the core (if the core is not radially laminated)
- hysteresis losses in steel structures (clamping frame, tank, core)

Eddy current losses in windings can be calculated very easily for 50 Hz or 60 Hz on the basis of the axial stray flux, or with the help of computer programs which take into account the distribution of the axial and radial flux density. According to simple theory, the eddy current losses of windings in watts ($Q_Z$) can be evaluated according to the formula:

$$Q_Z = K f^2 B^2 d^2 \lambda$$  \hspace{1cm} (W),

where:

$K$ is a constant
$f$ is the frequency
$B$ is the flux density (based on the load current)
$d$ is the conductor dimension normal to the flux direction
$\lambda$ is the conductivity of the conductor material.
For a two-winding transformer in connection with a six-pulse rectifier system, the load current in the secondary (LV) windings consists of a fundamental wave and harmonics of order 5, 7, 11, 13, 17, 19, 23, 25 and so on. If 120° rectangular current blocks are assumed, the size of the fundamental wave is 95.5% of the r.m.s. value, harmonics of order n decrease with factor 1/n in comparison with the fundamental wave. Under this assumption winding loss will increase rapidly, compared to a 50 Hz or 60 Hz load, especially if eddy current loss for the fundamental wave is high. If we assume an eddy current loss in the HV and LV windings of 10% (for only the fundamental wave) an increase in loss by a factor 1.7 would be the result if harmonics up to order 25 are taken into account.

In practice two effects alleviate this problem:
- the size of harmonic currents is smaller because current blocks are not rectangular and are longer than 120°, their shape will be influenced by the semiconductor control angle and the commutating reactance
- calculation of eddy current loss according to formula (1) is only valid for small conductor dimensions and low frequencies. If these conditions are not fulfilled, stray flux will be distorted within the conductors or, as a next step for large conductors in complete frequency range, eddy currents will reduce the stray flux due to the screen effect already explained for sheet or foil windings [14]. These details can only be evaluated by special computer programs
- additional losses in clamping structures, core, and tank caused by heavy currents along busbars, are reduced for higher frequencies by similar effects to those described before. Their calculation is very difficult, but loss measurements at high frequencies provide useful data [15,16].

Clearly, sheet or foil windings are more favourable in limiting loss increase with frequency.

For rectangular conductors it seems preferable to provide several parallel small conductors or transposed cables, to reduce losses due to harmonics. However, this solution would be contrary to the requirements in respect to dynamic behaviour unless the coils are cast into a solid block.

At this point it is important to note that for reactors magnetic flux in the windings is often a substantial part of the total magnetic energy. If the conductors have large dimensions (high current, hollow conductor with direct water cooling for instance) and if the frequency is high enough, magnetic flux in the region of the winding can be reduced by eddy currents. This means that the inductance will be different for d.c. and a.c. and in the latter case will depend on frequency.

5.3 Choice of connection for rectifier transformer

In principle for six-pulse rectifier systems with bridges, every three-phase two-winding transformer with any type of connection may be used. If 12-pulse systems are necessary in order to reduce the harmonic content of primary currents and ripple on the d.c. side, two secondary three-phase systems with a phase shift of 30° have to be provided. This can be achieved by:

1) transformers with two secondary windings, one delta-connected, the other Y-connected
2) two transformers with their primary windings delta connected, the secondary windings of the two transformers being delta and Y-connected respectively. Alternatively, the secondary windings can be connected identically with the primaries of the two transformers connected in delta and Y respectively.
3) two transformers with primary windings designed for ±15° phase shift respectively, both secondary windings being Y-connected.

The choice of these solutions depends on the requirements of the symmetry of no-load voltage and impedance between the different systems and also within the phases of the system.

In solution 1) two complete winding arrangements, consisting of a LV and HV winding, are placed on one core but separated in an axial direction by a defined space (called double-stack). To achieve good symmetry in the voltages, the ratio between the number of turns of LV1 (delta) and LV2 (Y) has to be very close to √3. Very few "couple numbers" fulfill this condition adequately, the smallest couple is 26/25. Other couples are 7/4, 14/8, 19/11, with increased tolerance in voltage ratio. With regard to the limited number of LV turns, optimized designs cannot always be used. There is always a certain amount of coupling between the two systems. The special problems of this double transformer solution will be discussed in section 5.4 but, from an economic point of view, it has the advantage of using only one core. Double-concentric solutions, i.e. a winding arrangement LV1-LV2, are not used.

Solution 2) provides total decoupling but requires two different transformers with different LV and HV designs respectively though the core would probably be the same. Manufacturing tolerances can increase design differences. By means of solution 3) this disadvantage can be avoided, since both transformers are identical in the mechanical sense, but with an HV winding in two parts to provide phase shift, usually as "extended delta". Nevertheless this proposal will fulfill requirements such as symmetry and decoupling in the best way. By means of phase shifts, 24 (or higher)-pulse systems can also be realized.

5.4 Special problems with double transformers

Double transformers, introduced in section 5.3 for 12-pulse systems, show some differences to the statements given in section 5.1 (regarding dynamic problems) and 5.2 (regarding the effect of harmonic currents) especially in connection with LV foil or sheet windings. Nevertheless, this type is used not only as the standard for 12-pulse rectifier systems in industrial service, but also in distribution systems (in this case both LV windings have the same winding connection) with rated power up to 90 MVA and rated voltage up to 110 kV. Nowadays cast-resin designs are more and more preferred.

In Fig. 6 the typical winding arrangement of a double transformer is illustrated. The two HV windings are connected in parallel. As a "black box", the transformer can be regarded as one three-winding transformer though, in reality, two 2-winding transformers are placed on one core.

The degree of decoupling between windings LV1 and LV2 is very important for use as a rectifier transformer. The decoupling factor, \( f_{dc} \), can be determined as follows:

\[
 f_{dc} = \frac{U_{LV10}\sqrt{3}}{U_{LV10}} 
\]

where:
$U_{LV10K2}$ is the no-load voltage on the terminals of LV1 while LV2 is short-circuited
$U_{LV10}$ is the no-load voltage on the terminals of LV1 while LV2 is in the no-load condition, with the same voltage on the HV terminal in both cases.

$f_{dc} = 1$ for two separate transformers.

The degree of decoupling depends on the distance, $a$, (see Fig. 6) between both systems; the greater is the distance, the higher is $f_{dc}$. Also, the distance, $d$, between the HV and LV windings, which affects the impedance between HV and LV, has some influence on the decoupling. With increased $d$, $a$ has to be increased to achieve the same figure for $f_{dc}$. Axial unbalance forces, which occur if only one of the two LV windings is shorted, as well as additional loss by harmonics are strongly bound to $f_{dc}$. The greater is $f_{dc}$, the better with respect to forces and loss.

Going into more detail, we have to distinguish between a symmetrical short-circuit, i.e. both LV windings short-circuited at the same time, and the unsymmetrical case, i.e. a short-circuit of only one of the two LV windings. In the symmetrical case only symmetrical axial forces occur which mainly affect the spacers between the two systems (distance $a$). If only one LV winding is shorted, a small current, depending on $f_{dc}$, flows in the HV winding of the other transformer. This means that for the complete transformer an unbalance in ampere-turn distribution is established, so creating unbalance forces. The forces try to increase the unbalance and the clamping structure has to be designed to accommodate them. Sheet or foil windings for the LV coil cannot improve the situation as in the case of a two-winding transformer. The problems become even more complicated since stray flux from the system under short-circuit conditions penetrates into regions of the other LV sheet winding (near space $a$), with the result that eddy current losses are produced. The situation during commutation corresponds to a short-circuit condition.
Each system works as a six-pulse system, which means that harmonics of order 5, 7, 11, 13, 17, 19, 23, 25 and so on are produced in the load current. Both systems together form a 12-pulse unit, with the result that HV primary current will not contain harmonics of order 5 and 7. These two harmonics (but also other pairs of order $6K \pm 1$, where $K = 3, 5, 7$ and so on) must therefore have different signs in the two LV windings, i.e. we have to consider the transformer formed by primary winding LV1 and secondary winding LV2 for this case. The two HV windings have to be considered as being connected in series, and carry some current depending on the degree of decoupling. If we assume the best decoupling ($f_{dc} = 1$, two separate transformers) the ampere-turn distribution between the HV and LV windings of each system would be balanced. The current in the HV windings in the case where $f_{dc} < 1$ will be smaller. This special current distribution over the complete transformer results in large radial stray flux components, and, subsequently, additional losses in the windings, cores, tanks and clamping structures. If the LV coil uses sheet windings, the loss increase is much higher since the distribution of current density is free and currents will concentrate towards the region of distance $a$. According to the condition of minimum magnetic energy, current in the HV windings will be smaller than for the case of windings with rectangular conductors [17].

These effects can be simulated with special computer programs which can calculate current density distribution and HV currents under minimum energy conditions depending on frequency. Calculations can be checked by direct measurements. In addition it should be mentioned that the harmonics of order 11, 13, 23, 25 behave like the fundamental wave. The result of investigations and experience is that in the case of designs with a grade of decoupling $f_{dc} \geq 0.9$, increase of losses due to the different effects of harmonics will be moderate, assuming that the additional loss for the fundamental wave is small. To achieve a specified figure for $f_{dc}$ it is necessary to increase the distance $a$ between systems if sheet windings are used rather than in the case of rectangular conductors for the LV coils.

In the past double transformers with intermediate yokes between the systems were used to achieve the best decoupling. These days, economic designs very near to standard programs, and especially dry types, are favoured. It is easier and cheaper to increase the distance between systems to achieve an acceptable degree of decoupling than to decouple by means of intermediate yokes.

6. **CONCLUSIONS**

Transformers and reactors, especially in connection with power converters, provide a wide field of problems for engineers in technical departments, during the manufacturing process, quality control and testing.

Core manufacture, choice of winding design under dynamic aspects and influence of harmonics in currents, as well as advantages of modern cast resin dry type transformers, have been described in detail. Some of the theoretical problems are treated very well in the literature as the references at the end of this contribution show. But the experience of both the manufacturers and customers makes an important contribution to further development in this special field.
REFERENCES


TRANSUDCERS AND MEASUREMENT TECHNOLOGY

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ABSTRACT

The long term reproducibility of the output current is an important parameter for a magnet power supply. For design purposes it therefore is imperative to know how to interpret the specifications of the components involved and how to apply them properly. After a review of error sources of DA-converters, a brief explanation is given of the operating principles of a flux-compensating magnetic current sensor. This sensor has proven to be a superior substitute for the high-current shunt. Also it will be shown that the correct interconnection of the different devices is of the utmost importance.

1. INTRODUCTION

Magnet power supplies usually are current controlled in order to obtain a reproducible magnetic field. The long term accuracy, an important part of the PARD (Periodic And Random Deviations) figure, is determined by the quality of the so called “front end”. As shown in Fig. 1, the front end comprises the reference source, the current measuring device or current transducer and the error amplifier. In the past, power supply engineers had to design these components themselves due to the fact that these were not available on the market. Today, all components are readily available to the highest possible standards. However, it still takes a lot of understanding to interpret their specifications in the proper way. Particular care is required towards the figures of gain, linearity, noise, offset and bandwidth. Problems like thermal EMF's, component ageing and thermal unbalance play an important role in this respect. Also the correct interconnection of high stability components is bound to a number of, often difficult to imply, rules. Incorrect connection of cable screens may contribute to the ripple and noise content and thus degrade the PARD-figure.

![Fig. 1 Power supply front end](image)

2. ERROR DEFINITIONS AND SOURCES

The different types of errors are presented in Fig. 2.

Zero offset, Fig. 2.1, means that a constant value is added to the original signal. It can be either expressed in μV or in ppm of FS, both being an absolute quantity. Examples are:
The input offset voltage of operational amplifiers as a function of time and temperature
- Thermal-EMF's, as for instance caused by the lead material which is used in some glass to metal hermetic seals
- Voltage drop in the signal common due to spurious current

The gain error, Fig. 2.2, is defined as a deviation of the nominal transfer ratio, assumed the offset is eliminated. A gain error is normally expressed in ppm. It will cause a deviation which is a fraction of the measured quantity. Examples are:

- Resistor changes as a function of time and temperature
- Gain drift of amplifiers with inadequate open loop gain
- Reference source drift as a function of time and temperature
- Poor joints or lead resistance outside a Kelvin circuit

The linearity error, Fig. 2.3, is the peak value of the deviation between the best straight line and the real transfer function. Normally it is specified in ppm of full scale. The linearity error may be symmetrised at the cost of a larger gain error. Examples are:

- Thermal feedback, in particular in operational amplifiers
- Lack of open loop gain of operational amplifiers
- Self-heating of resistors

Differential linearity errors may occur in DA converters as irregular steps in the transfer function. If the derivative of the transfer function changes sign on part of the trajectory, the function is no longer monotonous.

Dynamic errors can develop due to lack of bandwidth or to an insufficient slewing rate. In high current shunts errors can develop due to skin effect, self-inductance and lagging temperature of the resistor element.

![Fig. 2 Error characterisation](image)

3. **DIGITAL TO ANALOG CONVERTERS**

The majority of DAC's consist of CMOS types, using voltage switching techniques in conjunction with a R-2R ladder. This process is generally used for 8- to 12-bit schemes. Further details will not be covered in this chapter. High resolution types make use of different techniques, which will be reviewed here. References [1-2] may be helpful for more information about the subject.
3.1 Typical high performance schemes

The typical layout of a hybrid type high precision D-A converter is given in Fig. 3.1. This class of converters uses current switching bipolar technology. By means of a R/2R ladder network a number of binary weighted current sources is obtained. The absolute value of the current is directly related to the reference voltage by the reference feedback amplifier. The differential switches either contribute their part of the reference current to the output amplifier, or divert it to the analog ground. Bipolar operation can be obtained by introducing an offset of 1/2 MSB as indicated in Fig. 3.1.

\[\text{CONNECT TO V}_{\text{ref}} \text{ FOR BIPOLAR OPERATION}\]

1)  

2)  

![Fig. 3 Current switching DAC](image)

In connection with this scheme some less familiar but important specifications are:

- The current mode output resistance. The current switches, although operated in common base mode, still have a finite output impedance. This means that the collector-base voltage and therefore the input voltage of the output amplifier must be as close to analog ground as possible. This is expressed by means of the compliance voltage. The latter voltage is a direct measure for the maximum allowable offset voltage of the output amplifier. In the case that an external output amplifier is applied, care should be taken that the virtual input voltage is within some 100 μV from the analog ground.

- The source capacitance. The collective collector-base capacitance of the switching transistors can be as much as 200 pF. This means that, especially for high bandwidth amplifiers, capacitive feedback compensation is required to prevent high frequency instability.

The linearity which can be obtained with a current switching scheme is shown in Fig. 3.2. In order to describe the linearity of a DAC two definitions are needed:

- Integral Nonlinearity (INL). As already mentioned in Section 2, this is the maximum deviation from a straight line between the endpoints.
Differential Nonlinearity (DNL). The latter describes the individual bit weighting errors. These will cause erroneous steps at the major carry transfers.

As long as the DNL is smaller than 1 LSB, the transfer function is monotonic. From Fig. 3.2 it can be seen that in that case the INL is smaller than 1/2 LSB. This is a very important relationship for the calibration of a DAC. When the major carries are adjusted to within 1 LSB, the linearity is also guaranteed. There are a few exceptions from this rule though:

- Bit interaction. Unbalance between the current source transistors can cause the injection of error voltages in the R-2R ladder.
- Lack of open loop gain of the output amplifier. As the gain of the average operational amplifier varies with temperature and output voltage, the open loop gain should be large enough to cope with this problem.
- Thermal feedback. An increase in output voltage will cause an increase in the power dissipated in the output amplifier and the feedback resistor. In a badly designed DAC this could cause drift in the resistance ladder.

The certification of a DAC therefore requires a complete verification of the INL. How this can be done will be treated later on in this chapter. The measuring of the INL is even more important for other DAC schemes.

Figure 4.1 shows the principle diagram of a so called segmented DAC. This type of DAC operates like a Kelvin Varley divider. The full scale range is divided into a number of subranges by tapping one of the resistors of a linear divider. Obviously this design is inherently monotonous. The INL however can exhibit large errors, as shown in Fig. 4.2.

3.2 Error budgeting

Magnet power supplies often require 18-bit resolution. This means 262,144 steps of 1 LSB or 3.8 ppm each. Referred to a full scale output voltage of 10 V (forma 9.999962 V) each step is 38 μV. DAC specifications often use these units alternately. Therefore it is recommended to rearrange all data in the form of an
error budget. As an example the data sheet of a typical 18 bit DAC is compiled into the following table. The columns 1) through 4) give the error in ppm of full scale under the following conditions:

1) After calibration  
2) Over a temperature range of 10 K  
3) Over a time span of 1000 hours  
4) For a supply voltage variation of 1%.

<table>
<thead>
<tr>
<th>Error in ppm of FS</th>
<th>1)</th>
<th>2)</th>
<th>3)</th>
<th>4)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Zero offset</td>
<td>0</td>
<td>5</td>
<td>2</td>
<td>2</td>
</tr>
<tr>
<td>Gain (reference excluded)</td>
<td>0</td>
<td>10</td>
<td>6</td>
<td>1</td>
</tr>
<tr>
<td>Internal reference</td>
<td>0</td>
<td>30</td>
<td>5</td>
<td>1</td>
</tr>
<tr>
<td>External reference</td>
<td>0</td>
<td>10</td>
<td>25</td>
<td>1</td>
</tr>
<tr>
<td>Differential nonlinearity</td>
<td>2</td>
<td>4</td>
<td>*</td>
<td>0.5</td>
</tr>
<tr>
<td>Integral nonlinearity</td>
<td>2</td>
<td>3</td>
<td>*</td>
<td>*</td>
</tr>
</tbody>
</table>

The asterisk means that the value is not specified. But, as even the highest grade film resistors show a shelf-life stability not much better than 10 ppm/1000 h at room temperature, it may be expected that the monotonicity is lost within one month after calibration. This figure however strongly depends on the average operating temperature. Therefore it is recommended to operate a DAC at a relatively low temperature.

There is a remarkable difference between the internal and the external reference source. Although more sensitive to temperature variation, the internal reference shows a much better long term stability. This, together with the foregoing, suggests that this particular DAC should be operated with its internal reference at a stabilised temperature of for instance 10 °C. This could for instance be realised by means of a Peltier element.

If the required settling speed necessitates the use of an external amplifier, great care has to be taken towards the compliance voltage. As stated earlier, the amplifier's summing point should be within some 100 μV from the analog ground. This implies that offset trimming should be effected as indicated in the amplifiers data sheet, and not by superimposing a correction voltage on one of the amplifiers inputs.

A proper error budget is an excellent tool for judging the overall stability of an equipment. When summing the errors the correlation between them must be considered. If not correlated, the RMS sum should be taken. If the results of succeeding calibrations are logged, ageing can be traced as a means to determine the required mean time between calibrations.

3.3 Deglitching

Glitches may occur at the major carries (01 - 1 to 10 - 0) due to small timing errors between the internal switches. For this reason it is important to enter a new value with minimum timing errors between the individual bits. Glitches in the current reference can cause large voltage excursions in case of an inductive load.
Figure 5.1 gives an example of a simple "track or hold" deglitching circuit. During the transfer between digital inputs the switch opens during a short time. Due to the feedback capacitor the amplifier holds the old value. Once the switch is closed again the output tracks the new value with a speed determined by the closed loop bandwidth of the circuit.

Figure 5.2 shows a so called rate limiter. The tracking speed is limited to such a value that glitches never can cause an error larger than 1 LSB. Rate limiters do have the advantage that voltage excursions are limited to a known value independent from whatever input change. Care should be taken however that gain and offset drift of the circuit do not degrade the overall stability of the equipment.

![Deglitching schemes](image)

**Fig. 5 Deglitching schemes**

3.4 **Function generators**

In view of the often required complex waveforms and also because of the mere fact that a DAC is available, function generators should be made digitally. Figure 6 shows the general lay-out. Depending on the type of microprocessor, output latches may be required in order to prevent excessive glitching. In general, local processing power is required in view of the required resolution and speed. Full scale ramping in 100 ms with an 18-bit DAC requires a new input value every 400 ns. A look-up table may be needed in case of more complex waveforms, for instance to realise second-order rounding off of the corners of a trapezoidal reference waveform.

![Function generator](image)

**Fig. 6 General concept of a function generator**

3.5 **Test and calibration**

Figure 7.1 shows a test set up which can be used for fast calibration of DAC's with proven integral linearity. A high precision digital voltmeter can indicate the difference between the present and last stored value. If this feature is used to calibrate the major carries, INL and DNL must be within specifications. The function generator can consist of a bank of manually controlled switches.
Figure 7.2 depicts a test set up which lends itself for automated test purposes. The reference DAC should be calibrated at regular intervals and be temperature stabilised. Under computer control a complete analysis of a DAC can be made in a few minutes. At a few points within the full range the DAC under test should be allowed to settle thermally in order to unveil possible thermal feedback problems.

Fig. 7 DAC test and calibration

4. **DC CURRENT TRANSFORMERS**

Since the existence of high precision DC current transformers the high current shunt has lost its popularity. This is due to its relatively low output voltage, the lack of bandwidth, the lack of isolation and the poor long term stability. The operating principles of the Zero-Flux DCCT will be explained in the next sections. Further details are given in the References [3–6].

4.1 **The magnetic integrator**

The magnetic integrator is one of the main building blocks of the Zero-Flux DCCT. Figure 8 shows the fundamental layout. The transformer shown here is composed of a single ring core with several windings. Assuming that the core is not saturated, the amplifier will counteract any induced voltage and so stabilize the magnetic flux. This implies that the output voltage $V_{out}$ is a perfect image of the current to be measured, $I_0$. The maximum obtainable bandwidth of this circuit is solely determined by the stray inductance between both secondary windings. In practice a bandwidth of 1 MHz can be realised.

Due to the amplifier's offset voltage the ring core will saturate eventually. The feedback loop is then opened and so the circuit ceases to operate. To provide DC feedback as well, a second core is required in order to sense the absolute value of the magnetic flux.

Fig. 8 Magnetic integrator
4.2 The magnetic modulator

The absolute value of the magnetic flux in an iron core can be sensed by determining the symmetry of its magnetizing current. Therefore an excitation voltage $U_e$, see Fig. 9, is applied to an auxiliary winding. The symmetry of the magnetizing current $I_m$ can be sensed in different ways. As shown in the figure, its second harmonic content is a measure for the magnetic balance of the core. Although second harmonic detection provides excellent stability, it requires rather complex band filters and also a synchronous rectifier.

![Diagram of magnetic modulator](image)

Fig. 9 Magnetic modulator

The core must be driven into saturation in order to prevent backlash due to remanent magnetism. This makes it possible to apply a relatively simple circuit for flux balance detection, the peak-to-peak detector as shown in Fig. 9. If the diodes are properly matched a stability of 50 $\mu$A/K (0.5 ppm / K at $I_p = 100$ A) can be obtained if the proper core material is chosen. Due to the sampling nature of the circuit its bandwidth is theoretically limited to a value corresponding to the excitation frequency. The latter's value lies between 50 and 100 Hz. Higher frequencies not only tend to affect the accuracy due to iron loss, but also cause the induced voltage in the main circuit to increase to an unacceptable level.

4.3 The zero-flux DCCT

The Zero-Flux DCCT combines the bandwidth of the magnetic integrator with the stability of the magnetic modulator. As shown in Fig. 10, a third core is added to compensate the voltage induced in the flux compensating winding and in the main circuit, i.e. the circuit of which the current is to be measured. The flux compensating winding encloses all cores. In this way the magnetic integrator is kept out of saturation. Thanks to the third core the maximum output voltage of the amplifier is solely determined by the voltage drop in the compensating winding and the burden resistor.

![Diagram of Zero-Flux DC current transformer](image)

Fig. 10 The Zero-Flux DC current transformer
4.4 The burden resistor

In view of the required precision, the burden resistor should have the lowest possible dissipation and thus voltage drop. In practice a nominal voltage drop of 1 V is a good compromise in view of the stability of currently available high precision operational amplifiers. Applications like this require true 4-pole resistors. Although some high performance resistors are readily available on the market, most types of Zero-Flux DCCT's employ a proprietary design.

Figure 11 shows the construction of such a high precision resistor. The nominal resistance value is 1 Ohm, which accounts for a power dissipation of 1 Watt at a nominal current of 1 Ampere. A wire of a low TC alloy is loosely supported within a case filled with a heat conducting compound. Mechanical stress otherwise would lead to unexpected temperature drift. The wire is processed in order to obtain zero TC at room temperature. A special coating improves the long term stability. A bandwidth of approximately 1 MHz is obtained by routing the voltage tap leads in such a way that the inductive voltage drop of the resistance wire is compensated by magnetic coupling. The voltage taps are mutually heat sunk in order to prevent thermal EMF developing between the resistance alloy and the copper output leads. The described current sensing resistor has a temperature stability of 0.5 ppm/K and a long term stability of 0.5 ppm/1000 h at nominal load.

![Fig. 11 Construction of the burden resistor](image)

4.5 The output amplifier

To provide a full scale output voltage of 10 V, a high precision differential output amplifier is applied. The circuit is depicted in Fig. 12. Sense wires are foreseen in order not to affect the accuracy due to voltage drop in the output leads. The circuit board layout ascertains that thermal EMF's due to the different alloys used for amplifier and resistor leads are balanced out. In this way an accuracy of 0.5 μV/K is obtained for the input offset voltage. The gain stability is determined by the tracking stability of the resistors, the latter being better than 0.5 ppm/K.

![Fig. 12 The output amplifier](image)
4.6 External magnetic fields

When a permanent magnet is brought close to the magnetic modulator, Hopkinson's law explains that a great deal of the total flux passes through that part of space which has the lowest magnetic resistance. This includes the part of the core which is in the direct vicinity of the magnet, see Fig. 13.1. This part of the core will thus saturate asymmetrically and cause the magnetic modulator to generate an erroneous output.

Identical faults can be caused by wrong busbar placement. This is illustrated in Fig. 13.2. Any busbar position can be decomposed into the ideal, that is infinitely long and exactly centered, busbar and an external one, carrying the same current in the opposite direction. The latter conductor has the same effect as the permanent magnet mentioned earlier.

The sensitivity for external magnetic fields can be greatly reduced by a magnetic screen. This screen is applied between the auxiliary windings and the common flux balance winding, as shown in Fig. 13.3. In this way the screen is not magnetised by the main current which otherwise would tend to saturate it. Despite the screen the DCCT stays fully operational. This can be easily understood by applying Maxwell's first law: Closed field lines may circumscribe a conductor anywhere in space.

Fig. 13 The influence of magnetic fields on a DCCT

4.7 Test and calibration

Figure 14.1 shows a typical bench test set up in order to determine the exact number of compensating turns. By means of a multicore cable an identical number of turns is applied to both the DCCT under test and a current compensator with binary encoded taps. When both modulators show simultaneous zero output at a given ratio of test currents, n1 and n2 must be equal.

The value of the burden resistor and the gain of the output amplifier are calibrated with the circuit of Fig. 14.2. If the correct operation of the magnetic modulator of the DCCT under test is verified separately, this setup may be used for test as well as calibration purposes. The method does not take into account the rare chance that the number of compensating turns has changed spontaneously, for instance by replacement of a core and coil assembly.

Figure 14.3 demonstrates the use of a reference DCCT for on site test and calibration. The main current is applied to both DCCT's. The output voltages are compared by means of a high precision digital ratiometer.
5. **HOW TO INTERCONNECT CIRCUITS AND DEVICES**

Circuit designers are often puzzled by the question how to interconnect analog and digital grounds. Also the connection of cable screens and the grounding of equipment remains a matter of – sometimes hefty – discussions. Still there exist some simple rules to circumvent the common pitfalls. The difficulty of the subject lies in the understanding how to comply with these rules in a practical situation. The rules cover a wide range of frequencies, from DC to those frequencies determining the susceptibility of the equipment to electromagnetic interference. The subject is covered in depth by the References [7–8].

Figure 15.1 shows how to connect a DA converter properly. The zero reference connection points of the precision components in question, including the digital ground of the DA converter, must be connected to a single common point (shown in the drawing by a fat line) which must be as close as possible to the components involved. This line also is the zero reference for the local HF decoupling capacitors of the power supply lines. No other components whatsoever may be connected above the incoming zero reference, the point marked "0" in the figure. By doing so, the current in the zero line between reference source and DAC is only the DC current of the reference diode. The voltage drop can be determined by taking into account that the average copper track has a resistance of 1 Ohm/m for a width of 0.5 mm. Due to the large temperature coefficient of copper, the stability versus temperature may be seriously affected by wrong circuit board layout.

In general the DCCT will be situated on a separate circuit board or in a separate crate. Figure 15.2 shows the proper layout of the wiring in that case. It is assumed that the DAC and error amplifier are on the same board. The error amplifier subtracts both signals, amplifies the result to a fault tolerant level and restores the ground reference voltage. The following reasons lead to a wiring layout as shown in the figure:

- In order to overcome DC voltage drop in signal lines and connectors, a 4-wire (Kelvin) connection is applied between DCCT and error amplifier

- It must be assumed that DC and LF error voltages between the individual reference grounds G1, G2 and G3 are inevitable. Whether and how these points are really connected to earth is therefore not relevant. For this reason a differential input must be applied for the error amplifier. All voltage errors are now seen as common mode signals and do not affect the accuracy. Note that the resistors connected to the positive input of the amplifier do not necessarily require the same precision as their counterparts

- Capacitively or magnetically induced HF voltages, for instance caused by the firing of thyristors, often give rise to problems as the common mode rejection of the error amplifier deteriorates at higher frequencies.
The susceptibility for HF noise can be greatly reduced by proper application of twisted, screened conductors.

Figure 15.3 shows the general thought behind the connection of cable screens. If a HF voltage exists between both grounds, the cable screen will conduct a current determined by its natural impedance of roughly 1 \( \mu \text{H/m} \). The HF voltage across the shield will be transformed however to the inner conductors, as the latter ones share their magnetic field with the shield. This means that high frequency voltages between both equipments will be compensated and do not contribute to the common mode voltage. For the average cable this transformer effect starts at 1 kHz. It can provide a noise suppression of more than 60 dB at high frequencies. The effect strongly depends upon the optical transparancy of the shield, which in turn determines the transfer impedance of the applied cable. For braided shields a close texture gives the lowest transfer impedance and thus the best HF rejection.

Also shown in Fig. 15.3 is a common-mode inductor. By winding the cable one or more turns onto a ferrite ring core, the noise reduction will start at a lower frequency. This makes it possible to optimise the overall Common-Mode Rejection Ratio of cable and differential amplifier.

![Fig. 15 Interconnection schemes](image-url)
REFERENCES


MAINS RELATED PROBLEMS FOR PARTICLE ACCELERATORS

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ABSTRACT
The extensive use of power converters in particle accelerators gives rise to a number of problems for the electricity distribution especially concerning harmonics and fast voltage variations. This paper gives a general introduction to these problems and describes methods for reducing disturbance. Examples are taken from the CERN power network with its unusually high concentration of rectifiers.

1. INTRODUCTION

A majority of subsystems in particle accelerators require a DC supply: magnets, RF sources, vacuum pumps, particle detectors, electronics of all kinds, etc. The DC power is in all cases produced by a power converter fed from the AC mains.

In the past, the large converters were often in the form of rotating machines but have, with the progress in semiconductor technology, been replaced entirely with solid-state rectifiers. Today's rectifiers obviously have better performance in speed, regulation and reliability but this is at the price of two principal disadvantages. The first one is harmonics, generated by the non-linear rectifiers, a problem never encountered with rotating machines. The second disadvantage is the difficulty in filtering large power variations. Passive electric filtering is very difficult indeed compared to the old mechanical flywheel, where immense energies could be stored. Superconducting energy storage may one day be able to take over this task, but until then the mains will suffer from the consequences of load variations.

The mains, seen from the load, can in general be modelled as a voltage source behind a complex impedance. The voltage of the source is adapted to the transmitted power level following technical-economic considerations. The voltage levels are standardised by IEC and the most commonly used for industrial distribution in continental Europe are 400V, 3.3 kV, 10 kV and 20 kV. The higher voltage levels (>50 kV) are normally only used for power transmission over medium to long distances.

2. NETWORK IMPEDANCE

2.1 Impedance at power frequency (50 Hz)

The source impedance of a power distribution network at 50 Hz is dominated by its inductive components. The greatest contribution comes from the transformers' leakage inductances, but the inductances in cables and overhead lines should not be neglected. The source impedance is normally named short-circuit impedance as it determines the short-circuit current. The short-circuit power, defined as the product of the nominal voltage and the short-circuit current, is the most frequently used parameter when describing the performance of a network.
The resistance in the transformer windings and in the cables are important for the losses and consequently for the inherent network damping. It is convenient to formulate the time constant for a point in a network:

\[ \tau = \frac{L}{R} \]

Typical network values at CERN are shown in the table below.

<table>
<thead>
<tr>
<th>Voltage level (kV)</th>
<th>Time constant (ms)</th>
<th>Short-circuit power (MVA)</th>
<th>Short-circuit current (kA)</th>
</tr>
</thead>
<tbody>
<tr>
<td>400</td>
<td>30-60</td>
<td>7000-20000</td>
<td>10-30</td>
</tr>
<tr>
<td>66</td>
<td>30-90</td>
<td>300-3000</td>
<td>2.5-25</td>
</tr>
<tr>
<td>18</td>
<td>2-100</td>
<td>50-600</td>
<td>1.5-20</td>
</tr>
<tr>
<td>3.3</td>
<td>10-30</td>
<td>30-100</td>
<td>5-30</td>
</tr>
<tr>
<td>0.4</td>
<td>3-20</td>
<td>10-30</td>
<td>15-50</td>
</tr>
</tbody>
</table>

Capacitances are in general small except for HV cable networks, where an earthed screen around each conductor is compulsory for safety reasons. The effect on network impedance can in general be neglected at 50 Hz.

2.2 Impedance at higher frequencies

At harmonic frequencies, 150 - 2500 Hz, a reasonably simple model of the network impedance can still be used. Most impedances mentioned above are sufficiently frequency independent but HV cable capacitances must be included in models.

3. DELIVERED VOLTAGE LEVEL

3.1 Transmission voltage level

The level of the delivered voltage at the local grid connection is far from stable. In HV transmission networks the voltage is allowed to float slowly within a 10 - 20% range. In local HV distribution networks the voltage varies slowly in a 5 - 10% range due to local load variations. The ideal voltage source in any model must therefore be considered as a variable for some calculations.

3.2 Load variation

A load current will create a voltage drop over the complex source impedance:

\[ U_L = U_S - I_L \ast Z_S \]

The load current can be considered to be composed of an active and reactive part, the former being in phase with the voltage and the latter 90° out of phase. Thus, in the first approximation, an active load (resistive) will cause a voltage drop out of phase with the source voltage and a reactive load (generally inductive) will cause a voltage drop in phase with the source voltage (see Fig. 1).

As noted above, the inductive part of the source impedance is an order of magnitude larger than the resistive part. It is therefore clear that controlling the reactive part of the load current is crucial to good voltage control.
The traditional thyristor-controlled rectifier always consumes reactive power, the magnitude being dependent on the ratio of the AC input voltage to the bridge and the DC load voltage employed. It is important to optimise the rectifier transformer to the DC load to minimise the reactive power consumption. On the other hand, it is necessary to provide a certain overvoltage to quickly drive the load current in an inductive load to its final value. The shorter the ramping time, the higher the reactive power consumption.

It should also be noted that the good response time of the rectifier to a load change will be reflected on the mains side as a fast voltage variation, if no measures are provided for compensation. The implications for an accelerator should then become clear. The higher the reactive power swing and the shorter the ramping time, the greater the problem of controlling the voltage.

3.3 Voltage control

In an HV transmission network (400 - 66 kV) the power flow and the voltage levels are controlled through a coordinated adjustment at the different generating centres. The active power flow is regulated through the speed i.e. the generator shaft input power. The reactive power flow and the voltage level is regulated through the overexcitation of the generators.

3.3.1 Tapchangers

The slowly "floating" network feeds local distribution networks through transformers equipped with on-load tap changers (OLTC). The device functions like a variable transformer with steps of 1-2%, a response time of 15 - 120 s and a lifetime of hundreds of thousands of operations. The resulting voltage stability is around ± 2%, but with a limited speed of response.
3.3.2 VAr generating systems

The only fast and efficient way to perform local voltage control is to remove the major source of voltage variations i.e. stabilise the reactive load power. For the purpose of voltage control, it could be stabilised at any level, but it is in general preferable to set the level close to zero. This will minimise circulating reactive power, decrease the losses and improve utilisation of the transformers (see Fig. 2).

![Diagram of reactive power compensation](image)

Fig. 2 Principle of reactive power compensation

To stabilise the reactive power a suitable control element is needed. In the past the synchronous alternator was often used, but is no longer favoured because of high investment and maintenance costs. The saturated reactor was another alternative, but is hardly considered in recent installations due to high cost. The development of high-voltage, high-current thyristors has provided an excellent means for continuous control of reactors.

It is also possible to control capacitors with thyristors, but only in a step-wise fashion, and the system is more complex and inherently not as robust as the thyristor controlled reactor (TCR).

Some different TCR configurations are possible depending on the voltage and power level:

1. The reactors and thyristors operate at the network voltage (Fig. 3).

![Diagram of TCR direct HV configuration](image)

Fig. 3 TCR direct HV configuration
2. The network voltage is transformed to a lower level at which the reactors and thyristors operate. This can also be implemented through a tertiary winding on a major transformer (Fig. 4a).

3. A special transformer is employed, where the reactance of the transformer is used instead of separate reactors. The secondary voltage level is optimised for the thyristors (Fig. 4b).

Alternative 1 presents advantages in many cases:

- No need for special transformers, which are increasingly expensive to manufacture.
- Upgrading easier and less costly if foreseen at the design stage.
- Separate functions means that the consequences in case of breakdown are limited and relatively easy to foresee.

3.3.3 Control system techniques

Some different approaches can be taken in controlling the VAr production:

- The busbar voltage is used as control signal. This gives the best voltage control. A secondary control loop must be used to coordinate VAr production with the tap-changer. Otherwise there is the risk of saturation of the VAr generator.

- The measured reactive power upstream of the busbars is used as the control signal. In this case the VAr system is not effective against external mains voltage variations. There is no risk for saturation of the VAr generator and traditional tap-changer control is possible. It is the only approach if parallel compensators are used.

- An open loop control is derived from the control of the accelerator cycle. This can be very effective, as the control system delay is eliminated. However, it is somewhat complicated to program the control signal, in particular if many variations exist in accelerator operating conditions.
4. **STATIC PERTURBATIONS**

Static perturbations are defined as any quasi-stationary deformation to the sinusoidal power frequency waveform.

In the stationary case, a Fourier analysis can be performed, providing a decomposition of the waveform in harmonics of the fundamental. The harmonic contents can often provide useful information about rectifier and network performance.

4.1 **Harmonics generation**

Harmonic currents are always generated when a non-linear load is connected to the mains. Typical non-linear elements are saturating magnetic circuits, diodes and thyristors.

4.1.1 **Harmonics generated by transformers**

The magnetising currents to most transformers have a high content of all the odd harmonics, the percentages being determined by the saturation characteristics. The triple- \( n \) harmonics (3, 6, 9, etc.) are all but eliminated in delta-connected transformer windings and are therefore negligible. The fifth harmonic should be verified, but higher ones can in general be neglected.

4.1.2 **Harmonics generated by rectifiers**

Complete descriptions of diode and thyristor rectifier functions are abundant in the literature[1]. The theoretical analysis of the harmonics in the AC input current to an ideal rectifier with a purely inductive load, indicate that the amplitude follows the law:

\[
I_n = I_1/n
\]

where \( n \) is the harmonic order.

An ideal three-phase, six-pulse rectifier is perfectly balanced and does thus not produce any even, nor any triple- \( n \) harmonics, leaving only the 6k \pm 1, \( k = 1, 2, 3 \) etc. Increasing the pulse number to twelve, gives great cancellation advantages for the low order harmonics in particular, and the remaining ones are 12k \pm 1, \( k = 1, 2, 3 \) etc. In practice however, the pattern is modified by the load not being a pure inductance, by any passive filter at the output of the rectifier bridge and by the primary commutation impedance. A useful empirical formula[2] for the estimation of characteristic harmonics is:

\[
I_n = I_1/(n - 5/n)^{1.2}
\]

The harmonics that are theoretically predictable for a given rectifier type are called the characteristic ones, all others are called non-characteristic.

The non-characteristic harmonics are very much dependent on the quality of components, the thyristor firing control system and the supply voltage balance. In the absence of measured data, an estimate of 10 - 20% of the \( I_1/n \) value is often used.
4.1.3 **Harmonics generated by a TCR**

The branch current waveform is given by the following formula:

\[
I(\phi) = \frac{u}{\omega L} \cdot (\sin \phi - \sin \phi_1); \quad 0 \leq \phi_1 \leq \frac{\pi}{2}
\]

\[
I_n = -\frac{2\theta}{\pi \cdot \omega L} \cdot \frac{1}{n} \left[ \frac{\sin (n+1) \cdot \phi_1}{n+1} + \frac{\sin (n-1) \cdot \phi_1}{n-1} \right]
\]

and is shown in Fig. 5.

![Diagram of TCR circuit and harmonics](image)

**Fig. 5 Branch current waveform**

The branch current harmonics are given by:

\[
I_n = -\frac{2\theta}{\pi \cdot \omega L} \cdot \frac{1}{n} \left[ \frac{\sin (n+1) \cdot \phi_1}{n+1} + \frac{\sin (n-1) \cdot \phi_1}{n-1} \right]
\]

The spectrum of harmonics in the TCR current is very different from that of a rectifier. The amplitude of a given harmonic is not linearly dependent on the fundamental current, and higher harmonics are less of a problem due to the "roundness" of the waveform. The amplitudes vs. delay angle for branch harmonics are shown in Fig. 6.
Fig. 6 TCR harmonic levels
4.1.4 Switch-mode power supplies (SMPS)

The standard SMPS configuration can be seen in Fig 7. It is clear that the input waveform is far from sinusoidal. In fact, with average values of the system parameters, the dominant part of the input current is the third harmonic.

![Fig. 7 Switch-mode power supply waveforms](image)

It should be noted that, if a large number of single-phase SMPS are used in a three-phase system, the traditional rule of current cancellation in the neutral is totally invalid and, if the neutral conductors are not properly calculated and protected, they will most likely be overloaded and damaged. Upstream transformers must be severely derated because of the high circulating third harmonic current[3].

It is possible to make the input current almost sinusoidal. A large input filter inductance will stretch the current pulse and consequently lower the harmonic contents. A more elegant method would be a double conversion, where the first stage would be controlled to draw a sinusoidal input current from the mains. The power factor could as well be adjusted to any desired value.

4.1.5 Mains harmonics

The external mains network has a certain harmonic distortion. Any local harmonic filtering may find itself helping to filter the external mains as well as filling its intended function.
4.2 Harmonics filtering

The harmonic voltage generated at any given frequency is equal to the product of the network impedance and the injected harmonic current, both taken at that same frequency. The logical and most common way to minimise harmonic voltage generation is to provide a low-impedance path for the harmonic current. In general a separate path is required for each harmonic. Figure 8 shows a common configuration for a harmonic filter. It must be realised that for each branch providing a series-resonance, a parallel-resonance is formed with the source impedance. A typical impedance diagram is shown in Fig. 9. It is of greatest importance to keep track of the parallel resonances, frequency and gain, due to the presence of uncharacteristic harmonics and the inherent poor damping at these frequencies.

![Fig. 8 Typical harmonic filter](image)

![Fig. 9 Typical harmonic filter impedance diagram](image)
The design procedure is as follows:

- Make an inventory of the characteristic harmonics.
- Make an algebraic sum for each harmonic. Any dispersion in phase angle will contribute to a safety margin.
- Set up a model of the network with a network analysis computer program. Place a controlled current generator at the point of each (cluster of) non-linear load(s).
- Calculate the harmonic voltage at all interesting network nodes by setting the controlled current generators to the appropriate amplitude and frequency. Repeat for all interesting harmonics, non-characteristic included.
- Decide whether these levels are acceptable or not. If not, harmonic filtering is necessary. Experience at CERN and elsewhere shows that, in a HV network, each harmonic should be less than 1% of the fundamental and the total harmonic distortion be less than 3%.
- In general the economics of power factor compensation provides enough capacitors to make a good filter.
- Calculate the maximum impedance allowable for each filter branch. Calculate the L, C and Q necessary to achieve this impedance value considering mains frequency tolerance, component tolerances and temperature coefficients.
- Introduce the filter(s) in the network model and recalculate harmonic voltages. Reiterate if necessary.

The method described will ensure good harmonic filtering with a fair safety margin. If the network is weak, i.e. the total rectifier power is above 5% of the network short-circuit power, the parallel resonances between the filter and the network impedance may cause problems. It may then be advisable to add filter branches for the non-characteristic harmonics 3 and possibly 2. This will reduce the network impedance at frequencies sometimes critical to thyristor firing control systems.

4.3 Other filtering

The characteristics of a distribution network make it very difficult and impractical to filter higher frequencies (> 100 kHz). Standards have therefore been laid down as to the maximum permissible emission of radio frequency interference (RFI) from equipment (VDE 0875). The standards define measurement methods and maximum levels.

5. DYNAMIC PERTURBATIONS

This category includes all transients with a duration of less than 0.5 sec. The most commonly occurring ones are provoked by lightning strikes and by energising transformers and capacitor banks (filters).

5.1 Lightning overvoltages

The lightning may strike on or close to high-voltage overhead lines and particularly if the line is not protected by a guard wire. The strike will provoke a flashover to earth from the phase being hit. The flashover in general occurs across an insulator at the pylon nearest to the strike. The arc thus formed will continue to burn after the energy from the lightning has been dissipated, being fed with the short-circuit current of the network. The circuit-breaker(s) at the end(s) of the line will open as soon as the protection reacts, normally after 60 - 100 ms. An automatic reclosure is in general attempted after a suitable delay, corresponding to the de-ionisation at the flashover point.
The circuit-breaker may open one phase only, if the breaker is equipped with auto-reclosure and if the sophistication of the breaker and the protection so permit. Single-phase opening is normally practised on national grid systems above 100 kV.

The result seen at a remote load point is first possibly a short (50-500 μs) but high overvoltage (2-3 times U) followed by a single- or bi-phase voltage drop of any amplitude down to zero for the arc duration (60-100 ms) as shown in Fig. 10 for a typical event. In a downstream HV cable network, the initial overvoltage transient is not propagated very far due to the transmission line characteristic and can thus be neglected.

![Fig. 10 Examples of mains voltage sags](image)

Since most of the events are single-phase, a distinct increase in disturbance immunity can be realised through the systematic use of three-phase rectifiers i.e. drawing the primary energy from all three phases instead of only one. This approach combined with an appropriate downstream filter for energy storage will ensure that the downstream converter has a continuous supply of energy.

5.2 Transformer inrush current

When a transformer is energised, a transient phenomena called inrush, often occurs. The magnitude and duration is determined by the instantaneous voltage at switch-on, the saturation characteristics of the magnetic circuit and the remanence from its previous operation. The peak current can range from 5 - 20 times the nominal RMS current and the decay is often governed by more than one time constant. Typical waveforms are shown in Fig. 11 where the dominant decaying DC component should be noted. The resulting mains voltage drop is directly proportional to the mains impedance.
Fig. 11 Typical inrush current waveforms at the 18kV level

In general, the short-circuit impedance of rectifier transformers is low, as it is coupled with the desirably low commutation impedance for the downstream rectifier elements. In the past, if no particular request was advanced, this inevitably led to a design with a high inrush current. With increasingly larger rectifiers, it has become more important to limit the amplitude through careful design and today a reasonable value seems to be 6 times nominal current. With the introduction of super low-loss iron and automated stacking techniques, the typical decay time has increased, due to the lack of damping.

An efficient way to avoid the inrush current altogether is to pre-energise the transformer before the main circuit-breaker closes. This can be achieved either with series resistors from the normal source to the primary winding, a back-feed to the transformer via a secondary or a special pre-energising winding. The arrangement will incur added cost, complexity and sequencing, as well as additional safety problems.

5.3 Capacitor bank inrush current

At the moment of energising a capacitor bank, the instantaneous current is only limited by the local network impedance and, if employed, current limiting reactors. In the worst case the voltage will drop to almost zero and recover with a damped oscillatory response over a number of mains cycles. The determining factors are the ratio capacitive reactive power/short-circuit power and the network damping at the oscillation frequency. The higher the load on the network, the higher the damping.

The oscillatory response can easily generate several cycles of 30-50% overvoltage. This would previously not have given rise to great concern, but in the last 10 years the SMPS have been introduced in large quantities. In these devices there are several components, and in particular semiconductors, that are more or less exposed to the mains input voltage. *These components must be very conservatively rated to withstand overvoltages of this kind.*

6. PROTECTION PROBLEMS

The basic problem of protection is to discriminate between abnormal overcurrents that are harmful to the protected component and transients that can occur during normal operation. The problem is often solved by the introduction of a time delay before trip, the delay corresponding to the duration of normal transients.
6.1 Protection of rectifier transformers

The protection of a rectifier transformer follows quite closely that of a normal distribution transformer i.e. overcurrent, earth fault current, Buchholz and overtemperature. All of these traditional protection methods work well when properly set-up.

Some attention should be paid to the detection of earth fault currents. On a HV network the earth current is often measured as the sum of the three phases. This method works well as long as the CTs do not saturate. It must be pointed out that the risk of saturation has increased drastically with the prolongation of inrush currents as described in Section 5.2. Saturation times of up to 500 ms(!) have been measured at CERN. The problem is altogether avoided if the earth current is measured with a large CT encompassing all three phases.

It should also be noted that the same problem will recur at all upstream levels of protection independently of the CT rated current. This is because the problem is caused by the current-time integral and not the absolute value of the current.

6.2 Protection of harmonic filters

A harmonic filter must be protected against the following events:

- Overvoltage
- Earth fault
- Overcurrent
- Capacitor failure
- Harmonic overload
- Too fast circuit-breaker reclosure
- Network switching problems

Several of these headings represent protection of the HV capacitor units, whose dielectric is the most critical element. It should be noted that even though the internal fuse is highly recommended, it is only designed to protect against internal failures. It does not protect against externally created overcurrents.

The overcurrent protection of harmonic filters requires careful calculation and simulation of the switch-on transient. The protection characteristic must be set closely to the normal switch-on transient. The filter must be fully discharged before being switched back on again. This can be accomplished with a combination of discharge resistors or voltage transformers and a time delay on the circuit-breaker ON command. The filter must be separated from the network if an upstream circuit-breaker opens, or else ferro-resonance type phenomena may occur between the filter and connected, unloaded transformers. The resultant overvoltages (and currents) can easily cause capacitor failures.
7. EXAMPLE: CERN SPS AND LEP NETWORKS

7.1 The SPS compensator

The SPS accelerator has a repetition rate of 14.6 s at its maximum energy of 450 GeV. The consumption of active and reactive power during the cycle are shown in the Fig. 12. The one-line diagram of the 18 kV network is shown in Fig. 13. The majority of the pulsed power is drawn by the magnet rectifiers.

![Fig. 12 SPS 400 GeV cycle: power requirement](image)

![Fig. 13 SPS 18 kV network configuration](image)

Without compensation, the voltage variation at the 18 kV level would be 20 %. Even at the 400 kV level, the variation was judged to be unacceptable and thorough investigations were performed before the project was accepted[4].
The 18 kV busbars are operated in three sections, each being fed from a separate transformer. Two sets of busbars are used to feed all the pulsed loads of the accelerator and the experimental areas. These busbars each have a harmonic filter of 92 MVAR and a saturated reactor compensator of 120 MVAr. The third set of busbars feed predominantly stable loads and has a harmonic filter of 24 MVAr.

The voltage fluctuation at 18 kV after compensation is less than a percent. It should be noted that special attention was given to the accelerator cycle and in particular to slow down the transitions between flat-tops and ramps. This avoids oscillations due to the finite speed of response of the saturated reactor.

Figure 14 shows the harmonic distortion levels with and without filtering. The levels without filtering are only calculated as the accelerator cannot be operated without the filter.

![Graph showing voltage distortion at SPS 18 kV with and without harmonic filter]

Fig. 14 Voltage distortion at SPS 18 kV with and without harmonic filter

7.2 The LEP network

The LEP, being a storage ring, has a comparatively slow accelerating ramp and therefore does not create any problem for the mains network outside CERN. The one-line diagram is shown in Fig. 15. Most of the rectifier power is concentrated in points 2 and 6 and in consequence they were equipped with harmonic filters and TCR compensators. The harmonic distortion levels in point 2, with and without filtering, are shown in Fig. 16.

The 18 kV cable going around the machine in the 27 km accelerator tunnel gave rise to particular consideration. It is fed from the SPS "stable" network, which, as mentioned above, is equipped with a harmonic filter. When plotting network impedance at different points around the loop, it was noticed that the harmonic filter looses its efficiency with the distance, as the apparent series resonance frequency is moving away from the harmonic frequency it is supposed to filter (see Fig. 17). Local capacitor banks were added in points 4 and 8 to compensate the power factor and to improve filtering.
Fig. 16 Harmonic distortion at LEP Point 2

Fig. 17 Network impedance: LEP SE5
8. ACKNOWLEDGEMENTS

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PROTECTION, INTERLOCK, AND PERSONNEL SAFETY

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ABSTRACT
This paper covers the protection aspects for power converters, and includes a
detailed description of a conventional thyristor-controlled supply. Some particular
forms of protection are described, and the design criteria for high reliability are
given.

1. INTRODUCTION
As power converters range from a few hundred watts to several megawatts, it is difficult to establish
general rules for their protection.

For a low-power converter, a simple fuse and a unipolar circuit-breaker are adequate. In this paper
I will describe the protection needed for medium-to-high-power converters. The description is based on
the experience gained at the CERN Proton Synchrotron (PS) accelerator complex (Fig. 1) with about
1500 power converters of different types.

The protection aspects are very important and should already be taken into consideration at the
design stage. Protection added during construction or after installation of the power converter will
always be expensive.

After a general overview of the constraints that a power converter has to support, I will describe in
more detail the protection of a conventional line-commutated thyristor converter of 2.5 MVA used for
the Antiproton Collector (ACOL) main bending magnets.

Fig. 1 PS accelerator complex
Detailed descriptions already exist for the large converters such as the Main Booster and the Main PS power converter, and the protection has been adapted to the particular case (e.g. a motor-generator set with flywheel for the Main PS power converter). These methods of protection will not be described here, but details of the studies may be found in Refs. [1] and [2]. I will also mention briefly the protection of particular loads such as superconducting magnets.

2. GENERAL ASPECTS OF PROTECTION

The power converter transforms the incoming 50 Hz three-phase or single-phase current into a d.c. or pulsed current, thus satisfying the requirements specified by the magnet designer. The converter is inserted between the mains and the load: it has to cope with the mains on one side and with the demand of the load on the other side. The energy usually flows from the mains to the load. Only in the case of a pulsed current does the energy stored in the magnet have to flow back to the mains. In this case the converter has to be reversible in energy flow. The energy passing through the power converter will produce losses, and these losses will create heat.

2.1 Heat problems

The losses in the power converter are not equally distributed. So the heat which is produced has to be evacuated from the points where the losses are concentrated. This depends on how the current in the load is regulated. In converters with diode-bridges and transistor series regulators, the main losses will be concentrated in the regulator. As the efficiency of these converters is low, new designs employ switching devices instead of linear devices to regulate the output current. Only in particular cases are transistor series regulators still in use, e.g. for

- low-power application, such as power amplifiers with high bandwidth;
- correction elements that have to work near to zero or to cross zero smoothly;
- power converters feeding loads with very low time-constants, such as septum magnets.

In all other cases with output powers higher than 50 kVA, conventional thyristor-controlled converters are the rule. Here the losses will be concentrated in the thyristors.

In order to evacuate these concentrated losses, heat sinks are needed which transfer the heat to the surrounding medium. Apart from the highly concentrated heat production, all elements that carry currents will also produce losses. The magnetic elements will also have losses, which are induced by the varying magnetic field in the iron.

The protection elements against overvoltages, composed of resistor-condenser (RC) networks, will also have additional losses which have to be evacuated.

All this heat has to be extracted from the power converter, either by air or by water.
- If air-cooling is foreseen, the building has to be conceived accordingly, with false floor and an adequate ventilation system. In the power converter itself, either free air convection or forced ventilation has to be installed.
- If water cooling is foreseen, it is preferable to install a dedicated closed-loop water system in which the temperature and the conductivity of the water can be controlled.
- It is possible to have an intermediate system using indirect cooling with an air-to-water heat exchanger, or an oil-to-water heat exchanger when the elements to be cooled are immersed in oil.
This aspect of losses and of the evacuation of the generated heat pushes the designer to build power converters that have very high efficiency. This is the case for thyristor-controlled rectifiers where the output current is regulated by controlling the firing angle of the thyristor. One drawback of this solution is the bad cos φ of the primary current and the associated reactive power.

For medium- and low-range power converters, more and more switched-mode types are being used, which have high efficiency and a good power factor.

2.2 Electrical galvanic insulation

The power converter has to perform the galvanic insulation between the general voltage distribution and the load.

This insulation is needed in order to protect the load from earth currents in case of a fault in the magnet insulation. It is often necessary to fix the potential of the magnet to earth. In the case where several power converters are connected in series, the insulation of each converter permits a reduced voltage-to-earth of the magnet.

This insulation is satisfied by the rectifier transformer. The insulation of the transformer is of primary importance.

In the case of switching power converters, the insulation is performed by the high-frequency transformer.

If the load does not accept capacitive coupling with the mains, a screen has to be foreseen in the transformer between the primary and secondary windings. The screen will be connected to earth and will divert the capacitive current to earth. If no screen is installed in the transformer, these capacitive currents will affect the load. In order to reduce their influence, an additional high-frequency filter is normally connected to the converter output after the passive filter.

2.3 Incoming mains perturbations

The power converter has to be protected from the incoming mains overvoltage, even if this is of short duration, since the semiconductors used to control the output current are very sensitive to overvoltages. In order to block overvoltages with small energy-content, RC networks—so-called snubbers—on the semiconductor will be sufficient. In the case of overvoltages with higher energy-content, surge suppressors or spark gaps have to be foreseen. In general, these means of protection against voltage surges of high energy content are already foreseen on the voltage distribution side, and the power converter has to protect itself from the overvoltages induced by the switching actions in the converter itself.

2.4 Perturbation produced by the converter itself

The mains sees the commutation of the current in the converter as a short circuit. This short circuit will induce notches in the feeding mains with a frequency corresponding to the number of commutations: 300 Hz for a six-pulse circuit and 600 Hz for a twelve-pulse circuit. The amplitude of the notches depends only on the ratio of short-circuit impedance of the converter transformer to the short-circuit impedance of the feeding mains. If a high-power converter is connected to a weak feeding mains, then the perturbations created by this high-power converter can disturb other converters to the point that they will no longer work correctly. This is particularly true in the case of converters with thyristor control on the primary. In order to reduce this perturbation to a supportable limit, protection with LCR must
be foreseen. This protection will also reduce the incoming mains perturbation. It is also possible to incorporate this protection in a filter adjusted to the harmonic contents of the current. The high-frequency perturbations have to be filtered in order to meet the recommendations of the International Electrotechnical Commission (IEC) and the national standards.

2.5 Mains failures

The power converter has to be protected in case a phase is missing or if the voltage drops to a very low value. This is also of concern for the associated electronics. Thus it must be foreseen to keep the auxiliary voltage alive long enough to take the necessary steps to protect the power converter. The most critical part will be the control of the commutating thyristors.

2.6 Protection of the load

In general, the loads on power converters in an accelerator surrounding are magnets with high time-constants. The current in these magnets cannot increase or decrease rapidly without using high voltages. In case it is necessary to pulse the load current, the power converter has to provide a voltage that can be several times the ohmic voltage drop of the load. For a d.c. current power converter, an overvoltage of the order of 10 to 20\% of the ohmic voltage drop is foreseen. Therefore, the possible overcurrent in the load will be of the same order. For pulsed supplies, a special protection must be envisaged in order to limit the overcurrent. For the magnet to be de-energized more quickly than its time constant, it is necessary to have power convertors that are reversible in voltage and therefore return the energy to the mains. In the case of pulsed loads, this is very important since the mains has to accept this energy reversal. If the pulse rate is high, this may be unacceptable for the feeding mains. As an example, the main PS power converter power swing of about 80 MVA cannot be absorbed by the feeding grid and therefore a motor generator with a flywheel had to be inserted between the grid and the power converter.

In the special case where the load is a superconducting magnet, the converter may not be able to extract the current quickly enough in the event of a quench, and a special means of protection has to be studied.

2.7 Personnel protection

In addition to the protection needed for the equipment, a more important aspect is the safety of the personnel who have to work on the power converters.

From the point of view of personnel protection, the rules governing the construction of power converters are the same as those for other electrical equipment, and this aspect has to be taken into consideration already at the construction stage.

Electrical equipment is considered safe for personnel if no accidental contact with any live or potentially live component can be made. This is achieved by enclosing power converters in metallic cubicles which are well earthed. Construction is made according to the CERN Safety Codes, to IEC recommendations, and to the regulations of the Union Technique de l'Electricité (UTE) and the Electrotechnical Association. A special safety group at CERN keeps the codes up to date, publishes a Safety Code which is mandatory for all equipment installed on the CERN site, and inspects the installation to ensure that it meets the safety requirements [3, 4].
Particular care is necessary in the construction of large converters where the components cannot fit into cubicles. This equipment must be installed in separate enclosures with adequate protection. Access to such an installation is limited to skilled, instructed personnel only.

The general precautions that have to be taken when working with electrical equipment are as follows (see electrical circuit in Fig. 2). It is mandatory to
i) disconnect the feeding line;
ii) check that the voltage is off (the neon indicators are not a sufficient indication that the voltage is not on);
iii) take precautions to prevent the current being inadvertently switched on (put a lock on the feeder switch);
iv) mark the disconnected apparatus with a standard notice giving the name of the person who has made the disconnection, and the date and time when this was done;
v) discharge the installation, and in the case of power capacitors being installed in the equipment, close the discharge switches;
vii) never work alone.

![Electrical Circuit Diagram]

Fig. 2 Safety precautions
3. INTERNAL PROTECTION OF A CONVENTIONAL THYRISTOR-CONTROLLED RECTIFIER

Power converters above some tens of kVA are normally of the thyristor line-commutated type. In order to describe the protection of such a device in more detail, the 2 MVA power converter feeding the bending magnets of ACOL (Fig. 3) will be taken as an example.

For this power level, the medium voltage of 18 kV used at CERN is appropriate for reducing the current in the transformer primary. As the converter is too big to fit into one cabinet, it is split into three parts: i) the main circuit breaker, ii) the rectifier transformer, and iii) the converter itself, with a passive filter and a polarity reversal switch. In the following, the different items that perform the protection and the interlock of this converter will be described.

3.1 Mains circuit-breaker

The mains circuit-breaker is the most important device of all the protection elements: it has to disconnect the converter from the mains in case of an internal fault, even under the most difficult conditions; it has also to be adapted to the short-circuit power of the mains; and it has to protect the power converter against overload in case the internal electronic protection has not worked properly or was not adjusted to the right value.

For the power supply, we consider that the mains circuit-breaker has also to perform the task of switching on and off. Circuit breakers are not well adapted to this task because of their limited number of operations (about 10,000 at no load). Therefore, for supplies with smaller ratings, connected to the 380 V a.c. three-phase system, the task of protection and the switching actions are separate. The switching action is performed by the main contactor.

At the 18 kV distribution level, only breakers are used. A regular check must be made to ensure that the number of operations is not exceeded.

At the moment of switching on, there will be a high current peak (inrush current). The inrush current depends on
- the magnetic core material;
- the transformer construction, especially the short-circuit impedances;
- the applied voltage and the allowed iron losses;
- the line impedances;
- the instant of switching on with respect to the mains voltages; the worst moment is during the zero crossing of the input voltage.

Table 1 shows approximate maximum values for the inrush current, compared with the transformer power ratings [5].

<table>
<thead>
<tr>
<th>Nominal power of the transformer (kVA)</th>
<th>250</th>
<th>500</th>
<th>1000</th>
<th>2000</th>
</tr>
</thead>
<tbody>
<tr>
<td>Ratio of $I_{peak}$/inrush to nominal current $I_n$</td>
<td>20</td>
<td>15</td>
<td>12</td>
<td>10</td>
</tr>
</tbody>
</table>

The amplitude of the current peak will decay to half the value after 10 to 20 periods. For higher power ratings, the decay will be slower. The circuit breaker must not react to this high peak current.
Therefore, during the switching on, there must be a delay before opening. Also, this high inrush current disturbs other users on the grid; therefore, whenever possible, inrush-current suppression should be foreseen.

3.2 Inrush-current suppression

In order to limit the current during switching on, the transformer is connected to the mains via current-limiting resistors. After a delay, the main contactor will short-circuit these resistors and the transformer is thus connected to the full mains voltage.

By choosing convenient values for these protection resistors and by having a delay before the action of the main contactor, it is possible to avoid producing any overcurrent at switching on. The following photographs were taken on a 12-pulse 20 kVA power converter:

Plate A without inrush-current suppression;
Plate B with inrush-current suppression;
Plate C nominal current of the power converter.

Plate A:
Without inrush current protection;
vert. scale I = 60 A/div.;
hor. scale time = 20 ms/div.

Plate B:
With inrush current protection;
vert. scale I = 60 A/div.;
hor. scale time = 20 ms/div.

Plate C:
Power converter at nominal current;
vert. scale I = 60 A/div.;
hor. scale time = 20 ms/div.

The values of the inrush-current-limiting resistor and its power ratings are not easy to determine. An empirical rule is to assume that, at switching on, the transformer is fully saturated and that the current is limited by the resistor only.

If the current is limited to 2.5 to 5 times the magnetization current of the transformer, then the transient will decay in less than five periods (100 ms). After this delay, the vector diagram of the circuit will correspond to the values given in Fig. 4. The lag \( \phi \) between the feeding voltage \( U \) and the voltage on the transformer is of the order of 10° elec., and no saturation of the main reactance of the transformer will appear when the main contactor C1 short-circuits the resistors after about 100 ms.

The power rating of the resistors has to be chosen so that they can support several switchings on in one minute. If the main contactor will not switch on, then the resistors have to support the full power for an indefinite time. In order to reduce the ratings, an adequate thermal protection acting on the auxiliary contactor C2 must be foreseen, according to the cabling diagram shown in Fig. 4.
3.3 Power transformer protection

The construction of the power transformer is a special task, which has been described in another paper given at this conference. For medium-power converters, free air convection is usually adequate for cooling. The transformer has to be adapted for rectifier operation, which means that there will be square current waves with high harmonic content, and corresponding high losses due to the skin effects in the windings. The maximum hot-spot temperatures that are allowable for safe operation are determined by the type of insulating material. In order to ensure a long lifetime for the transformer, the allowed temperature rise is fixed a class lower, and thermostats for this class have to be fitted to the windings. The allowed hot-spot temperature should not be higher than 120 °C, with an ambient temperature of 40 °C.
The magnetic circuit has to be foreseen for continuous operation at a nominal voltage of +10%, so as to reduce iron losses.

High-voltage transformers (18 kV on the primary) are usually cooled by oil. We still have an important number of PCB-cooled transformers with ‘Buchholz’ protection. According to Swiss regulations, they must be replaced by transformers without PCB. We are investigating the use of moulded-resin rectifier transformers as a replacement. For transformers that are installed indoors, silicon oil is normally used as the insulating and cooling medium. In certain cases the silicon oil is cooled in a special oil/water heat-exchanger. The corresponding interlocks for the water flow and the oil temperature have to be foreseen.

3.4 Overvoltage protection against voltage transients

Capacitor–resistor networks are the most common form of surge protection. If connected across each of the incoming lines, they will protect the rectifier from high transients. Instead of capacitor–resistor networks, it is also possible to use ‘selenium’ surge suppressors, voltage-dependent resistors (MOV), or spark gaps to prevent voltage surges. A useful technique is to connect a low-current rectifier bridge across the incoming line, with a series condenser–resistor network on the d.c. output terminals of the bridge. This is known as a ‘bucket’ circuit [6]. The advantage of this circuit is that a d.c. capacitor can be used to reduce the dimensions of the protection. The rating of this protection circuit depends on the transformer ratings and on the stored energy at switch-off. A reasonable value for the capacitors is about 10 to 30 μF. Fuses are used to protect the rectifier bridge against short circuiting.

3.5 Overvoltage induced by the hole-storage effect at the end of commutation

It is necessary to have condenser–resistor networks across each thyristor, because the hole-storage generates voltage transients at the end of commutation.

The highest level of overvoltage occurs when the end of commutations coincides with the peak of the mains voltages.

The design of the components of the capacitor–resistor network depends on the following parameters (see Fig. 5):
- transformer reactance;
- the inverse peak current, which depends on the di/dt and the forward current before commutation;
- the stored charge in the semiconductors. The stored charge of the thyristors is a given parameter, which is indicated in the manufacturer’s data sheet.

The practice is to limit the overvoltage to less than twice the peak supply voltage. This is a good compromise between the thyristor voltage rating and the size of the capacitors. The usual safety factor for the thyristor is 2.5 to 3 times the peak supply voltage.

The optimum design is normally intended to minimize the value of the capacitance and the power dissipated in the suppression resistor. The resistor must also limit the discharge current of the capacitor when the thyristor is switched on. In the case where several thyristors are connected in parallel with anode reactors to equalize the individual current, the first conducting thyristor will take over the current of the capacitor–resistor protection of the other parallel thyristor. The value of this current should not be higher than the allowed instantaneous current when the thyristor starts conducting (∼ 100 A for converter power thyristors).
3.6 Fuses as overcurrent protection for the thyristor rectifier

Because of the low heat capacity, the junction of a semiconductor device is very quickly damaged by large current overload. Thus, semiconductor circuits are usually protected from such an overload by the use of fuses or by combinations of fuses and circuit breakers, all of which must be carefully selected. Fuses have to be selected according to the following criteria:

- **Nominal current of the device to be protected.** Fuses react on r.m.s. values, thyristors react to peak currents.
- **P1t:** the P1t let-through of the fuse has to be compared with the P1t of the semiconductor. The P1t of the fuse should always be less than that of the device.
- **Overload current ratings.**
- **Fuse overvoltage ratings,** when a fuse clears; the resultant dI/dt produces a voltage transient. This arcing voltage should match the system overvoltage.

For very high currents, the losses inside the fuse will increase the temperature of the fuse itself; thus the current rating of the fuse changes. Care must therefore be taken to ensure that the temperature at the end-caps of the fuse is maintained at a reasonable value.

The usual way to protect the rectifier against overload would be to place the fuse at the d.c. output as shown in Fig. 6. Since the load of the rectifier is a magnet with high inductance, it is not possible to have the fuse in this position. The fuse could not clear the fault. The protection of the load has to be done by other means. If several semiconductors are connected in parallel, the only way to introduce the fuse would be in position (b). The fuse can then clear the fault and the current in the load is not interrupted. The most convenient place for the fuse is in position (c), the feeding branches of the bridges.
In case the output current is too high to be taken by one individual thyristor per branch, it is better to parallel a full bridge assembly and not the individual thyristors.

3.7 Passive filter protection

The magnetic elements of the filter will be protected from overheating by means of thermocontacts fixed on the coils. The filtering capacitors and the damping resistors are protected by fuses. A filter capacitor can carry only a reduced amount of a.c. current. The series inductance of the filter has to be such that it will reduce this current to well within the limits of the allowed ripple current.

The fuse cannot be dimensioned according to this current since, at switching on, the charging current of the filter is very high. The fuse has to be large enough to support this peak charging current:

\[ I_{\text{peak}} \approx \frac{U_{\text{step}}}{\sqrt{L/C}} \]

If the rectifier is of the unipolar type, an electrolytic capacitor may be used.

The power rating of the damping resistors could be reduced by using an r.m.s. filter-current protection. An alternative solution would be to protect the damping resistors of the filter by means of a thermocontact fixed on the resistors itself, which switches off the converter in case of overheating.

3.8 Earth protection

The load and the converter cannot be left floating; somehow they have to be connected to ground. If the electrical insulation is damaged, or if the load or the supply is inadvertently connected to earth, then
current will flow to earth and may build up to a too high value. Therefore adequate earth protection is needed.

If the magnet has not to be fixed to earth potential at one end, then the best solution is to have a differential medium-point earthing on the supply, as shown in Fig. 7c. By so doing, the voltage-to-earth of the magnet coils will be only half the maximum d.c. value of the converter voltage. For large accelerators with hundreds of magnets connected in series, it is in any case necessary to split the power converter into several subunits, with each unit left floating. Only one unit will be earthed at one point with a low-value resistor. The voltage on this resistor will be monitored and used as an earth-current monitor. If this current exceeds a certain value, it will be detected and the power converter will be switched off.

For a lower voltage converter, one output terminal may be earthed according to Fig. 7a or 7b. The connection to earth as in Fig. 7a is the simplest and does not need special adjustment.

**Fig. 7 Earth protection**
If the converter has to be completely floating, then a detector, as shown in Fig. 7c, may be convenient. Such a differential detector is commonly used in a.c. distribution systems for the protection of the personnel.

4. INTERNAL INTERLOCK

There is no clear distinction between faults and interlock. I will make use of the following definitions:

Interlocks are indicators that some action has to be taken before the converter is ready to be put into operation.

Faults result from exceeding the protection levels, or they are caused by the breakdown of components.

4.1 Door interlock and short-circuiting of capacitors

If the converter voltage is higher than 50 V or the stored energy is higher than 10 J, then the converter has to be switched off and made safe, and the condenser has to be discharged and short-circuited to earth before any potentially live parts of the converter can be touched.

To make sure that these rules are followed, the cubicles containing the apparatus are equipped with door contacts that switch off the power supply if the doors are opened; or the cubicle is surrounded by panels that require special tools to be opened.

According to French safety rules, it is also necessary to earth the incoming mains before work can be carried out inside a cubicle. This earthing has to be done with a cable, of adequate cross-section, which is fixed to the incoming terminal by means of studs. Special care has to be taken with high-voltage converters. The circuit breakers are of the fully encapsulated type. Work can be carried out on the equipment only if the breaker is withdrawn and the line shorted to earth (standard 18 kV circuit breaker at CERN).

For particular equipment such as the modulator power converter in the preinjector of the CERN LEP, it is foreseen to have permanently installed earthing poles. An interlock allows the mains to be switched on only if all earth connections are fixed at their open position. This is to avoid switching on when part of the equipment is still short-circuited or earthed.

4.2 Air-cooling interlocks

4.2.1 Free-air convection

No special interlocks are needed for equipment working with free air convection. The protection given by the thermocontacts placed on the elements to be protected is adequate. In general, the components are specified to work with a maximum ambient air temperature of 40 °C.

It has happened that during hot summers the temperature in the equipment building rose to very high values, and exceeded 50 °C at the air inlets of the power converter. Since it was not possible to keep the apparatus running under such conditions, a system of forced ventilation had to be installed. In certain buildings it was even necessary to introduce a heat exchanger in order to refrigerate the incoming air to an acceptable value, and also to keep the accelerators working under these conditions. In new buildings, a dedicated ventilation system is always installed.
If such an installation is foreseen right from the start, then it is possible to have the building temperature-controlled. This has great advantages with respect to drifts in the converter current, i.e. day and night variations can be reduced.

In the converter itself, the elements that could generate heat are protected by thermostats placed at the hottest spot. For semiconductors, it is convenient to place several thermostats on different cooling fins in case cooling ducts become blocked or individual thyristors take more current than others (missing firing pulses, imbalance of the current in parallel-connected devices). In the case where several thermocontacts are connected in series, the contacts should be equipped with indicators (which can be reset), in order to identify the element that has tripped.

Should it not be possible to extract the heat by convection, forced-air cooling may be foreseen.

### 4.2.2 Forced-air ventilation

The temperature increase in a piece of equipment can be considerably reduced by fans, which makes it possible to increase the current-carrying capacity of semiconductors to a great extent. Nevertheless, ventilators have a high fault-rate owing to

- wear on the bearing of the shaft;
- overheating of the motor if the voltage is not adapted (at CERN, the incoming phase voltage is often 220 V + 10% permanently).

In addition, high-speed ventilators are also a source of noise. Low-speed ventilators are more reliable and less noisy, but, since the air speed is reduced, the efficiency of the cooling is also reduced.

In any case, when forced ventilation is used, adequate protection of the fan motor by means of the associated interlock has to be envisaged. In addition, airflow detectors are currently used: either a flap with a microswitch; or small turbines with speed monitors; or heated temperature-dependent resistances with their own detector. The fans have to be inspected regularly and eventually greased.

From the foregoing it is clear that it is preferable to employ forced ventilation only when the available space does not permit another solution to be used, and where water cooling is not possible.

If the converter losses are too high to be evacuated by air, then water cooling is preferred.

### 4.3 Water-cooling interlocks

If water is to be introduced in a power converter, the designer has to take this into consideration at the earliest possible stage. For equivalent power rating, preference must be given to a solution with higher currents and lower voltages.

Electrical insulation has to be assured by demineralized water and by insulated water-tubes of a length sufficient to give the necessary impedance. The possible overpressure in the system has to be checked, and pressure-reduction valves may have to be used. The diameter of the cooling hole has to be large enough to avoid blockage. Filters may have to be inserted. The temperature of the complete water-pipe system has to be kept above the dew-point level in order to avoid condensation.

If the converter is installed in a building where the temperature can fall below 0 °C, then there must be the possibility to empty the water system.

These requirements are best fulfilled with the aid of a separate, closed, demineralized water system, and with a heat exchanger. Such a system is convenient only for several power converters or for a single high-power converter of several megawatts rating.
The demineralized water will eliminate the leakage currents and prevent erosion of the metallic parts of the water circuit. It is nevertheless necessary to use only stainless steel and copper for the metallic part (no brass). A temperature regulator will easily keep the water system above the dew-point level.

If these requirements are satisfied, then a water-cooling system will prove to be as reliable as an air-convection system. The amount of ventilation in the building can be reduced, since the losses are partially evacuated by the water. For the power converter to work, the following interlocks have to be provided:
- a flowmeter, with an auxiliary contact if the water flow is adequate, installed at the water outlet of the power converter;
- a thermostat near the water outlet.

4.4 Interlock for the polarity changeover

If the power converter is equipped with a polarity inverter, then the polarity changeover switch should not be activated before the load current has disappeared. Should this switch open before the current has reached a very low value, then the voltage induced by the $\text{d}I/\text{d}t$ on the load inductance will build up to such high values that the insulation of the magnet coil will be damaged if no spark gaps are foreseen on the magnet terminal. The contacts of the polarity switch itself will also be damaged, and if the load current is still high when the contacts open, then the damage caused may be very important.

It is imperative to have a safe procedure for the polarity inversion system. The following sequence will be adequate:
- The changeover switch should not change position during switch-on or switch-off of the converter; therefore, use a motor-controlled changeover in preference to inversion contactors.
- Before the changeover switch is activated, decrease the current to zero.
- After a delay, the current should be really zero.
- If it is not, then switch off the main contactor.
- Wait some seconds (depending on the time constant of the load), then start the polarity changeover.

![Diagram of zero-current detector](image)

**Fig. 8** Zero-current detector
In order to control that the current is really zero, a dedicated zero-current detector has to be foreseen. This may consist (Fig. 8) of a) a shunt with a current detector; b) a DCCT with incorporated zero-current signalization; c) back-to-back diodes. The parallel resistor prevents the system from being too sensitive. The remnant current in the load is 0.7 V, divided by the resistor values. A drawback of this solution is the high losses on the diodes for high-current power converters. For moderate currents, it is a very simple and safe detector (the diode has to carry the possible overcurrent of the load).

For fast current inversion, a mechanical inverter will not be adequate, so thyristor inverters have to be used. For a current crossing from positive to negative, or vice versa, without interruption, an antiparallel power converter with circulating current is the best solution. The ratings of the two converters may be different, but if equal current is needed in both polarities, then the two power converters must be identical. In order to limit the circulating current, limiting chokes have to be inserted in series with each converter.

5. EXTERNAL INTERLOCKS

5.1 Magnet interlocks

In case of electrical insulation damage of the magnet coils or a short of the magnet terminals to earth, internal earthing protection of the power converter will be adequate. The current limitations of the converter may not be enough if the converter is not matched with the load, or if the cooling water for the magnet is switched off. Thermocontacts, fixed to the magnet coils, protect these coils against overheating. In order to interrupt the current before the trip temperature of the thermocontact is reached, the magnets are usually equipped with a flowmeter at the water outlet of the cooling system. Should the flow of water be interrupted, this device will provoke a shutdown of the power converter. Some magnets may also be equipped with a water-pressure monitor as an extra means of protection. Normally, a red push-button fitted on the magnet is used to switch off the converter in case of emergency. The same button is also used to prevent the magnet from being switched on should work have to be carried out on the magnet itself. Cables, of adequate insulation and cross-section, are used to interconnect the individual magnets. Usually, the magnets have covers over their busbars, and as long as this protective cover remains in place, no live electrical parts can be touched. Should the cover be removed, its interlock contacts will cause the power converter to be switched off. This protection system allows people to work in the experimental areas or around the accelerators, even if the magnets are energized.

The connection between the magnet interlock contacts and the power converter has to be very reliable, as these contacts are the only means of protection against overheating in case of water-flow interruption or overcurrent in the magnet.

When changing the magnet to a different power converter, care must be taken to ensure that the corresponding interlock has also been changed. This is even more critical if one magnet is powered by several power converters of the same type or of different types, e.g. capacitor discharge and pulsed d.c. power converter [7].

If one magnet or a series of magnets are pulsed by several power converters, also in series, then the interlock of each magnet has to be able to switch off all the series converters.

The use of several interlocks of different types (thermostat, water-flow supervision) is enough to produce a certain redundancy of magnet protection.
5.2 Protection for superconducting magnets

The problem of protecting superconducting magnets is very vast, so here I will only sketch the main points that have to be taken into consideration when designing power converters for such loads. The time constant of a superconducting magnet, including its cables, is very high. In order to build up the current, only a limited voltage is available if the feeding converter is not overdimensioned in voltage ratings. In case the power converter trips, or if the mains voltage disappears, the only voltage available for reducing the current to zero is the voltage drop in the last conducting semiconductors and the voltage drop on the cables and connections to the magnet. It may take several minutes or up to hours before the current is down to zero. The semiconductors must be able to carry this current during all this time without any external cooling. If the converter needs water or forced ventilation in order to work normally, it must be foreseen to have a separate discharge element that can work without any active cooling in case of a fault on the power converter (for example, free-wheeling diodes). The other problem of superconducting magnets is the loss of the superconductivity, called 'quench'. When a magnet quenches, current has to be extracted from the magnet in order to reduce the magnet stress to a minimum. In Fig. 9 a possible quench protection is shown. If the quench detectors see a asymmetry in the voltage across the coils, then a fast d.c. circuit breaker is actioned, and the energy stored in the superconducting coil is dissipated in the external discharging resistor. In order to de-excite the magnet by means of the power converter in the normal way and to return the energy to the feeding mains, the diode of the discharging circuit can be replaced by thyristors.

![Supramagnet quench protection diagram](image)

**Fig. 9** Supramagnet quench protection

5.3 Safety interlocks

Some equipment buildings are closed down and no access is allowed when the magnets are energized. This is the case when the magnetic fields are very high, the zone is exposed to radiation when the accelerator is working, or the busbars of the magnets are not protected. In order to ensure personnel protection, access is allowed only when the power converters are turned off. This interlock system is usually integrated in a more complex general access system, which includes the condition that all power converters be turned-off to give access to the zone.
6. PHILOSOPHY CONCERNING OVERLOAD AND DEVICE FAILURE

6.1 Thyristor fuses

Our experience with the more than 1500 power converters on the CERN PS site, which range from several hundred watts to 80 megawatts peak power (PS main supply) and which include a considerable variety of types and ages (some dating back to the sixties), has shown that the thyristor is a very safe device if correctly protected against overvoltage. It is very rare that a thyristor breaks down. Therefore, for a medium power supply up to about 250 kVA we have suppressed the thyristor fuse as foreseen in subsection 3.6. This is valid only if capsule-type thyristors are used, because when submitted to overloads that are several times higher than the I\textsuperscript{2}t capacity of the thyristor, this type of thyristor will present a short circuit instead of breaking up, thus provoking an electric arc.

6.2 Faulty thyristor

A thyristor fault nevertheless cannot be excluded; in this case, fast-acting detection and action on the primary main contactor is needed. The integration of such overcurrent protection is shown in Fig. 10. The detector threshold is adjusted slightly above the nominal current of the power converter. Since the inrush-current suppression has eliminated the need to delay the switching action of the contactor, the fault can be cleared immediately. Eventually, the thyristor that is carrying the short-circuit current in series with the faulty thyristor will also break down.

![Fig. 10 Thyristor fault](image)

6.3 Overcurrent

The primary current is measured by a.c. current transformers. If overcurrent is detected, the action is twofold: the firing pulses are pushed to full inversion and the main contactor is opened. If one thyristor
is already defective at switch-on, the overcurrent will be kept within reasonable limits by the soft release of the firing pulses out of inversion.

6.4 Protection of power converters in case of pulsed d.c. load

In pulsed converters the available voltage at the output of the supply allows a load current to build up, which can be several times the nominal peak current of the load. For protection, a distinction must be made between the different currents:
- peak current: $I_{\text{peak}}$
- r.m.s. current: $I_{\text{rms}}$
- mean current: $I_{\text{mean}}$

In general, the thyristor rectifiers are designed to conduct the nominal flat-top current, which is also the peak current of the load. The cooling for the semiconductor depends on the mean current. The magnetic elements, the cabling, the fuses, and the interconnections should be designed for the r.m.s. current. For accelerators, the peak current corresponds to the maximum field and therefore to the maximum particle energy. Often after the commissioning of the accelerator, an upgrading in energy is needed. This means working at a higher flat-top current. This is possible for the power converter as long as the r.m.s. current value is not exceeded. For the semiconductors this is possible if the nominal current was the flat-top current. The flat-top length has to be shortened and eventually the cooling has to be improved.

An important aspect of programmed power converters is the protection of the power equipment in case of faulty reference programs. This can be done by testing the reference on a simulator before it can be applied to the converter. The elaborated magnet current cycles are then fixed in tables. If the programme changes too often, this is not convenient, and special action has to be taken in order to ensure proper functioning of the electronics.

In a power converter used for magnetic measurements where fuses had to be replaced too often because of wrong setting of the current reference, a clever means of protection was found: each fuse was fitted with a thermocontact which acted as an interlock on the power converter before the r.m.s. current of the fuse was reached. The $I_{\text{peak}}$ was set so that the thyristors were well protected, and a thermocontact on the thyristor heat-sinks prevented overloads.

7. Grading of Protection

It is possible to distinguish between passive and active protection.

Passive protection is given by the safety factors that are chosen at the design levels. It is more convenient to speak about the reliability and the lifetime of the individual components of a power converter. The aim is to use well-specified elements with known ratings and lifetime. The individual components should be tested.

Active protection is given by all interlock systems and fault-detection devices installed in the power converter to prevent faults and malfunctioning of the converter.

An important task of the converter designer is to make a good compromise between active protection and safety margins, and to reduce the protection to a necessary minimum. For example, it is better to choose thyristors with higher current ratings than to introduce a complex r.m.s. current protection.

A certain degree of redundancy in the protection devices has to be foreseen: e.g. a water-flow monitor and a thermostat; overload protection on the main circuit breaker and current limiting by the current regulation system.
With the advent of microprocessors installed in the electronics of the power converter, it is now possible to have better supervision and to monitor the faults. This is very convenient for the operation and the preventive maintenance of power converters.

7.1 Signalization

The status of the converter has to be indicated. All interlocks should also be indicated so that action can be taken to reset these interlocks and put the converter in operation. Once the power converter is operative, no fault should appear. Some faults will not trip the converter at once, but only after a given delay:

- water flow too low,
- filter current too high,
- current imbalance in a 12-pulse system,
- current regulator saturated.

These faults may disappear after a short time, so they will not interrupt the supply of current to the load. Some indicators can transmit a warning to the crew operating the power converter:

- transformer temperature above the first threshold,
- timing signal outside some defined values,
- current setting outside some predetermined settings.

When several elements are connected in parallel and the converter is designed according to the principle of n + 1, operation can continue even if an element fails. The associated fuse will clear the fault, and a signalization tells the personnel to replace the faulty elements when the converter is stopped for servicing.

Mains perturbation:

i) The mains perturbation is so important that the output current is outside the precision requirement.

A signalization with no action on the converter has to be provided. Possibly an event counter can be foreseen, in order to keep a record.

ii) One phase or all phases disappear. If the converter is connected to the mains by a contactor, then the voltage drop will eventually be so long that the contactor will trip and the converter will stop. In order to keep the control of the converter active for a certain length of time, the auxiliary supplies for the electronics and for the interlocks have to be supported by electrolytic capacitors which should be able to bridge short interruptions of the mains up to 20 ms.

For the main power converter of the accelerator, the auxiliary supplies for the interlocks and the electronics are derived from uninterruptible power supplies that can handle interruptions of the mains lasting several minutes.

7.2 Faults and interlocks that trip the converter

The interruption of the converter current may be achieved by

- setting the reference to zero,
- inverting the thyristor rectifier,
- tripping the main contactor.

The first two actions correspond to a Standby status of the power converter; the main contactor or circuit breaker in the OFF position corresponds to the OFF status.
In normal action the following time sequence should be respected:

i) Setting the reference to zero. This will produce a decay of the current in the load according to the programmed rate-limit of the reference signal.

ii) Inversion of the thyristors.

iii) Actioning of the main contactor when the current is zero. The switching ON procedure should be done in the inverse sequence, taking into account the premagnetizing of the transformer if it is implemented.

By so doing, contact wear on the main contactor will be reduced, and the number of operations of the main circuit breaker of the contactor will depend only on the mechanical limitation of the switching actions. In addition, stress on the semiconductors is reduced.

Certain faults or interlocks have to act directly on the main contactors or the circuit breaker and have to be backed up by the control electronics. This allows the fault sequence to be traced back, and the origin of the first fault to be found.

The faults that should trip the main circuit breaker are:

- an internal fault on the incoming line or on the transformer;
- a fault in the thyristor rectifier;
- thyristor fuses blown;
- primary overcurrent;
- a door interlock;
- an earth protection trip;
- loss of control of the converter, e.g. interruption of a feed-back signal, oscillations;
- a fault on the control electronics, e.g. loss of the auxiliary voltage.

All other faults should follow the normal sequence: first Standby then OFF.

7.3 Crowbar circuit

Some power converters are equipped with a thyristor crowbar, backed up by a mechanical crowbar. These protective devices are only of interest in special cases such as very high load-inductance, and if in this case the power converter loses all control action, or the feeding mains fails, or if a very high voltage surge on the magnet terminals will be generated.

7.4 Tripping of the high-voltage feeding line

General emergency stops are installed in each rectifier building or experimental area. In case of increased risk conditions (testing of the power converters, installation of new equipment, checking of the functions on working equipment or installations), a special ‘cut-off’ emergency stop has to be provided which will switch off the equipment under test or the whole area in which work is going on. This kind of check has to be carried out by skilled and trained personnel only (reserved for people with an electrical operative licence).

8. FIRE PROTECTION

In power converters, there is a very high concentration of power, especially when forced ventilation is used, and because of this high concentration, hot spots develop inside part of the converter. If the power source is not cut off, the temperature will rise to such a degree that there is risk of fire. The usual way to protect the individual elements from reaching such high temperature levels is to install thermocontacts, which will trip the power converter.
8.1 Fire protection for power transformers

Figure 11 shows a power converter after a fire. This fire was caused by a transformer that was made up of copper bands. As the insulation between the bands was damaged, a short circuit developed in the adjacent copper band. This short circuit affected only a small area of the copper, and a hot spot was created. The short-circuit current from the feeding line was too low to trip the overload protection on the primary. The thermocontact did not react, so the short circuit continued to be fed until a fire broke out, resulting in very serious damage to the converter.

![Damage caused by a fire in a power transformer](image)

Fig. 11 Damage caused by a fire in a power transformer

In order to prevent such mishaps, all power converters with this type of transformer were equipped with a special fire-detector cable [10–12]. The cable consists of two twisted-pair conductors covered with thermally sensitive insulation material. If the temperature increases above a given value (cables exist for temperatures ranging from 85 °C to 105 °C), the insulation will melt and the two conductors will be shorted. This short is translated into an interlock that will trip the supply. If insulating materials that do not propagate fire are used, then the fire will die out. This kind of protection has worked well. Manufacturers have developed an improved design for new transformers using better insulation material, thus making this kind of fire protection obsolete [12].

8.2 Fire detectors in the equipment buildings

Other hot spots may build up in a power converter, and the risk of fire is still present. In the converter, many cables or busbars are connected together by bolts and screws. If a connection loosens, then the contact resistance increases. The losses in the connection can increase the temperature to the point where corrosion builds up. After some time the connection can melt, causing an electric arc and setting fire to the converter. Careful inspection of the converter at the commissioning stage, and good maintenance, will reduce the probability of such occurrences.

In order to limit the spread of fire and to reduce subsequent damage, all power-converter buildings and all experimental areas and equipment rooms at CERN have a smoke- and fire-detector installation.
The signalization of all these detectors is transmitted directly to the Fire Brigade control centre. When an alarm is triggered, the firemen go quickly to the scene of the fire and endeavour to bring it under control. There are also many false alarms that call for intervention by the firemen. However, thanks to their prompt action, no serious fire has broken out during recent years.

The following is an example of an intervention:

In the summer of 1989, the main circuit breaker of a power converter did not switch on properly, but the signalization showed that the supply was on. The power transformer was connected to the 380 V through the limiting resistors of the inrush-current protection. After a certain time the resistor became so hot that it started to smoke and triggered the smoke detector.

The surrounding components of the resistors melted but no fire developed, and the damage was very limited because of the rapid action of the firemen.

9. CONCLUSIONS

The new generation of power converters are very safe because of
- the high individual rating of the semiconductors; no need to connect several thyristors in parallel or in series;
- new material for the electrical insulation: flame-retardant, with low production of corrosive gas and smoke in case of a fire;
- guaranteed long-life condensers (computer grade), able to sustain high peak currents and ripple current;
- transformers designed for rectifier applications;
- the use of cables without PVC in power-converter installations.

Because of all these precautions, the protection aspect has become easier. In addition, both the control and the interlock based on mechanical relays are more and more frequently replaced by an extremely reliable integrated system. Thus, once the questions of electrical noise and interaction are settled, these devices will work without any problems of intermittent contacts and mechanical wear.

The use of microprocessors for the supervision of power converters allows easy diagnosis in case of a fault.

The trend towards the use of convection cooling and dry transformers increases the reliability still further.

It is important to use well-designed and dependable electronics that have been tested and burnt-in before installation.

The protective action has to be tested and adjusted at regular intervals so that it will work when needed and not give false alarms. After all, it must not be forgotten that no device or piece of equipment exists which has an indefinite lifetime and which will never fail.
REFERENCES

REMOTE CONTROL AND FAULT DIAGNOSTICS

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ABSTRACT
Interfacing power converters to an accelerator control system presents the designer with a number of special considerations. The all-important initial choices are explained and some of the major pitfalls highlighted. The way in which noise can affect performance is frequently emphasised. Digital/Analog methods are outlined as are certain practical considerations.

1. INTRODUCTION

All particle accelerators built since 1970 have relied on computer control of their power converters, so much so that today it would be unthinkable to operate an accelerator without remote control of these essential components. Today, the vastly increased size of accelerators plus the increase in the number and variety of power converters have dictated a much more systematic design approach. Failure to appreciate the particular problems involved in interfacing power converters to the accelerator control system can seriously compromise both performance and reliability. It is important to keep in mind that a wide variety of disciplines, technologies and power levels are involved and must interact successfully to produce the degree of performance required by modern accelerators. This is reflected in the following sections which are developed as five main themes namely:

i) system design requirements;
iit) analog signal treatment;
iiii) digital signal treatment;
iv) fault diagnostics;
v) application software.

2. SCOPE

Unfortunately the scope of the subject is extremely large and as a consequence it is impossible to give more than an overview in certain areas. The main aim is to give power conversion specialists and users of power converters a deeper insight of the requirements, problem areas and solutions so that these can be kept in mind during the design stage and subsequent exploitation.

3. THE ACCELERATOR ENVIRONMENT

It is frequently assumed that the environment in an accelerator is less harsh than in industry. While some conditions, such as dust, corrosive atmosphere etc. are not normally a problem in accelerators, other conditions such as electrical disturbances on the mains network, radio frequency interference, electrical noise, temperature variations etc. are altogether comparable with, or exceed industrial level.
As a consequence, the correct physical placement of power converters in an accelerator complex is vital if both high performance and a robust remote control are to be obtained. This important initial consideration is developed in the next section.

Detailed knowledge of the local environment in which power converters will operate is therefore a prerequisite to system design. Poor siting of converters and/or incomplete evaluation of ambient conditions e.g. placement of converters in the accelerator tunnel, have in the past been major causes of difficulty.

4. **SYSTEM DESIGN, COMMISSIONING AND MAINTENANCE**

The design of the remote control system of an accelerator is a highly complex task and is in constant evolution. It is not intended to propose a particular approach here, but more to point out that the control of the many hundreds of power converters typically found in accelerators should have a profound influence on the overall design.

Essentially, the remote control of a power converter can be broken down into the following areas (Fig. 1):

i) operations control room work-stations, displays etc;
ii) data transmission to equipment clusters, timing etc;
iii) interconnection of power equipment into clusters;
iv) interfacing of digital data to the power converter;
v) interconnection of the power converter to its load, power source(s), machine interlocks etc.

![Diagram of System Areas](image)

**Fig. 1 System Areas**

This breakdown could of course apply to most other accelerator equipment. The particular area of interest is the interfacing of digital data to the power converter, and it is this aspect which will be treated in detail.
The starting point therefore must be an accurate specification of the required performance of each power converter plus full specifications of all material (hardware and software) which will interact with this interface. This is a non-negligible but vital task. Operator requirements are frequently unknown at the start of a project, performance definitions are sketchy but usually have a tendency to require higher stability, more resolution etc. Perhaps the most dramatic change to foresee early in the design study is the extension of the total number of converters and/or a fundamental change in method of interconnection or operation. This means that the approach MUST be extensible to cater for variable numbers of converters and wherever possible that changes can be made in software rather than in hardware. All this suggests that a programmable interface is highly desirable.

Where performance requirements of the power converters are tighter than 5.10E-4 (the majority of accelerators), the local environmental conditions play an important role. Detailed knowledge of the following is necessary:

i) expected mains network voltage variations, including short term transients, drop-outs etc;

ii) local ambient temperature variations, both short and long term. Due care should be taken to evaluate not only the temperature rise in the building due to power converter dissipation, but also the self-heating in the interface electronics crate. The use of air conditioning equipment and/or local fan cooling can be critical if air velocities exceed 0.5 m/s, due to the "chill factor";

iii) humidity variations, particularly if local condensation can occur, e.g. water-cooled equipment;

iv) proximity to other equipment producing RF or other electrical or magnetic fields;

v) placement of cables, where power and signal cables must be well separated. In this respect, due attention must be given to the total length of all cables. Particular care should be exercised where cables carry very high currents (kilo-amps).

The design study can then continue with the analysis of analog precision requirements i.e., resolution, linearity and stability, suitably divided into control-loop errors, environmental effects, component drifts etc. This analysis usually results in a so-called "error budget", where the sum of all known error sources must be less than the required precision. An rms sum can be derived to indicate the typical performance and evaluate "trade-offs". In practice, a linear sum is to be preferred since the real errors are usually unknown, can vary over time or temperature, and the resulting pessimistic value gives a reasonable performance margin. Great care should be exercised if manufacturers' data sheet values are used for this analysis, since experience has shown such values to be often highly optimistic. A much preferred method is to measure realistic samples of critical components under the expected operating conditions. While this is initially time consuming, the benefits can be enormous both from a cost as well as an overall performance point of view. From this analysis, the performance of the reference source (usually a DAC) and the output measuring system (usually an ADC) can be extracted and suitable components specified.

It remains to quantify the approximate number of digital control signals which will be needed to interact with the complete power converter as well as a corresponding number of digital status signals. It is advisable to include some "spare" control and status signals for future expansion or modification.

The bringing into service, or commissioning, of the interface equipment can be a tedious process, relying as it does on the availability of all of the other material to which it is connected. The same conditions occur during repair and maintenance periods. It is therefore extremely useful to build test/simulation possibilities directly into the interface right from the design stage. Additionally, such tools can be of benefit when diagnosing fault conditions (see section 12).
5. **GENERAL HARDWARE CONSIDERATIONS**

5.1 **Overall Control System Standards**

It is normal practice to define certain hardware standards early in the design phase of a project. These will normally comprise recommendations for such items as racking, crates, cables etc. and modules for unifying the control communication, timing and computing aspects. Clearly such standards are vital in obtaining a coherent control system. However, attempts to use standard modules which are not correctly matched to the requirements of high performance power converters should be avoided. This is particularly true for DAC's and ADC's where both noise and grounding considerations are of paramount importance. The idea of grouping together many power converter interfaces in one common crate is therefore not to be recommended for very high performance systems, as this gives rise to unavoidable common-mode voltages between the single interface crate and the power converters. Such an approach compromises the overall performance and can also be difficult to repair during operation. The apparent economies resulting from such an approach are often lost in more complex physical interconnections and higher cabling costs. Whilst this approach has been used extensively in the past, it can create insurmountable problems in matching the wide variety of converter types and realisations to one standard interface. The result is frequently a very minimum control ability and restrictions to the evolution of the various parts of the power converter system.

5.2 **Connecting to the Power Components**

Connecting the power part to the control interface is a considerable problem, due to the wide variety of types of power converter. Every attempt must be made, none the less, to standardise the functions required. Some basic principles must be followed here, due to the level of interference normally present, namely:

i) place the electronics (protection/regulation and remote control interface) in a separate metal enclosure, preferably in a rack adjoining the power part(s);

ii) use isolation techniques for all digital signals, taking care of the peak voltages generated under fault conditions;

iii) ensure that only one point of both the protection/regulation and remote control interface is connected to the power converter ground;

iv) buffer all precision analog signals using differential techniques if they are not precisely related to the single ground point;

v) pay attention to all cabling into and out of the power part by routing cables away from magnetic components, contactors/relays, power cabling etc. Use short screened twisted-pair cables wherever possible to reduce pick-up. Group into separate cables those signals liable to mutually interfere with each other, e.g. firing pulses, digital control, digital status and analog. [A good knowledge of RF Interference problems and preventive methods can be most valuable in this area.]

5.3 **Components, Assembly and Testing**

A complete set of electronics for a modern power converter may contain more than 1000 electronic components. The choice of these components therefore will determine, in large part, the reliability of the complete converter. Equally, the quality of the assembly and the testing of each module is vital. To achieve the highest
possible system reliability it is recommended to underrate most components by about 30% with respect to their normal maximum conditions (voltage, current or temperature). Experience has shown that by paying due attention to these factors, (in particular quality control, burn-in, and extensive automated testing) power converters can demonstrate mean time between failures (MTBF) of more than 100,000 hours.

6. THE USER INTERFACE

The definition of the user interface is essentially a written list of "commands" describing the required functionality of the complete power converter. In an operational accelerator there are two distinct groups of personnel who interact with power converters, namely support staff and operations staff. These two groups have to be able to use the equipment in different ways to perform their necessary duties. Extensive use is made of the remote control environment by both groups, therefore as far as possible a common user interface should be provided. It is vital that for example, for setting the output current both groups use the same hardware and control methods, and not one uses binary while the other uses decimal. In this context the interface designer has the responsibility to evaluate carefully his choice of hardware/software so as to facilitate the overall use of the power converter system.

6.1 Support Personnel

Support staff must frequently test, calibrate and repair the overall system. The time needed to perform these tasks can be reduced by providing adequate facilities for this work. In most cases extensions of the "operational" commands can provide a good solution. A special case concerns software development, which is often an ongoing task. The design of the user interface must therefore take all these considerations into account in order to maximise system availability.

6.2 Operations Personnel

The requirements of operations staff have frequently been neglected in the past. However, a large amount of time is spent by operators in switching on, setting values, checking and adjusting power converters. It is essential therefore that these tasks are made consistent and are as easy as possible to perform. While most interaction is made via an application program and not directly with a power converter, the interface design plays an important role. At the moment, a considerable effort is being made in the accelerator community to unify control methods, with the aims of reducing the amount of specialised software needed. For these efforts to be successful an equal effort on the part of interface designers will be necessary. This aspect is expanded in section 13.

7. THE ANALOG/DIGITAL INTERFACE

7.1 Noise Considerations

Noise must be considered as one of the most important sources of error in high precision applications. Not only can the environmental noise be considerable but also the various noise sources inside the power converter cubicle can be some of the worst to counteract. Noise is generally defined as an unwanted AC signal superimposed on the wanted signal. Major AC noise sources are:
i) mains frequencies (50 Hz) and higher harmonics;
ii) commutation of power switches, thyristors, GTOs, MOSFETS;
iii) auxiliary switching power supplies;
iv) digital circuits, microprocessors, buses etc.;
v) spark generators, such as relays, circuit breakers etc.

The frequency components of these noise sources extend from very low frequencies up to several hundreds of megahertz (Fig. 2). Analog circuits can malfunction when subjected to noise, for example operational amplifiers which are used for many purposes in the power converter and its interface "see" these noise sources in series with the input signal and react in various ways according to the amplitude and frequency of the noise (Fig. 3). It is clear that signals lying within the limit of the op-amps bandwidth will be treated normally. However, out-of-band noise can be rectified by the op-amp and thus produce apparent DC offsets which vary both with the frequency and amplitude of the noise (Fig. 4).

It must also be remembered that the rejection of differential amplifiers drops rapidly with increasing frequency due to high frequency mismatch of the gain-determining components. This means that differential circuits cannot always provide adequate noise rejection for wide-band noise.

![Fig. 2 Frequency Spectrum of Noise Sources](image)

![Fig. 3 Noise Injection Areas in Typical Op-Amp Circuit](image)
Digital circuits can equally be affected by noise, although in general the level must be considerably higher than for analog circuits. Sequential logic is particularly vulnerable to impulsive noise (e.g. microprocessors).

A further consideration is the statistical nature of noise. This is frequently overlooked and is a cause for some concern when using analog to digital converters in particular. Noise is usually measured as an rms value and Fig. 5 shows the probability of various peak amplitudes occurring on a wanted signal in the presence of Gaussian noise. It can be seen that if noise is truly random in amplitude (which it rarely is) then there is a finite possibility of false readings occurring even when the rms noise amplitude is more than eight times smaller than the specified resolution. Therefore, making accurate measurements in the presence of considerable noise needs a very careful choice of method.

The increasing use of switch-mode power converters highlights these noise problems in all the above areas, so much so that some older designs, which operate successfully with thyristor equipment, will no longer perform correctly. Experience has shown that noise suppression techniques must be applied correctly both at the source as well as on the critical analog components (i.e. control loops, ADC's etc.).
7.2 Precision

The definition of precision when referred to power converters varies widely and can be a source of considerable misunderstanding between suppliers and users. It is preferable to define precision as the variation of the actual output compared to perfect performance. In practice it is convenient to split the definition into two parts:

i) requested output compared to actual output;
ii) measured output compared to actual output.

The actual output must be referred to certified international electrical standards (e.g. Amps). The variation should be expressed as parts per million (ppm) of the full range output (e.g. ±35 ppm of 1250 Amps). This method avoids the confusion arising when a precision is calculated at say 19% of the output range. Performing laboratory measurements to these levels of precision requires very expensive apparatus and should not be underestimated.

The requested output is a numerical value which passes via the communication mechanism to the interface. In the interface it is normally used to set a DAC which can only be set to discrete values defined by the required resolution. This results in a setting error of up to ±1/2 LSB. Apart from this setting error, the precision of the actual output value is given by the analog errors of the DAC, the feedback loop, and the output measuring transducer (Fig.6). In practice the DAC errors predominate.

![Diagram of DAC and feedback loop including error sources.](image)

**Fig. 6 Major Error Sources in a Typical Power Converter**

The measurement of the actual output value is somewhat more complicated depending on whether an additional output transducer and centralised/distributed digitising hardware is employed. These various options are shown in Fig. 7. The preferred configuration consists of an independent output transducer and an ADC. In this case the measuring precision is given by the analog errors of the transducer and the ADC plus the digitising error of the ADC (usually > ±1/2 LSB). It is frequently assumed that these errors are inferior to the above "total setting error" but in reality they are altogether comparable if not greater. This means that great care must be taken if values read from the ADC are subsequently used to set the DAC, since unless the total of all errors are much less than 1 LSB then cumulative errors will occur. The resultant effects on the long term reproducibility of the accelerator are both confusing and difficult to diagnose. It is therefore recommended to only use the output measuring system as a means of confirming that the required output is within a certain tolerance and not to use a "closed loop" method.
7.3 Digital to Analog Converters

A typical DAC consists of four essential parts as shown in Fig. 8 namely, a precision voltage reference source, n bit switches controlling binary weighted current sources and a summing output amplifier. The value of n determines the resolution of the DAC, for example where n = 16 each increase in the input binary value will increase the output by 15 ppm (65535 steps). Although other methods are possible, the above represents the current industrial approach. It is not intended to explain in detail how such a DAC operates as there are now many excellent books on the subject [1]. However, a thorough understanding of the performance limitations, sources of error and variations of parameters with both time and temperature is essential. Many industrial products were not designed to be used as ultra-stable DC reference sources but rather as variable AC sources (e.g. CD players, CRT deflection systems etc). It is for this reason that in general the temperature coefficients and long-term stability are not adequate. Only a few standard industrial products are capable of reaching a true 16 bit performance, although great strides have been made in the last two years. Every DAC should be checked for monotonicity, output temperature-coefficient and long term stability before being put into service.
Fig. 8 Essential Parts of a DAC

Most manufacturers of high performance DACs provide instructions on how to interconnect them with their associated components. It cannot be emphasised too strongly how important it is to avoid earthing currents and a connection scheme as shown in Fig. 9 is recommended. It is extremely easy to ruin the performance of a good product by not paying attention to circuit board layout, connections to the control loop and auxiliary power supply performance.

Fig. 9 DC Connection Scheme to Avoid Earthing Currents

Figure 10 shows a DAC which is located at some distance from the control loop. The use of an isolation or instrumentation amplifier becomes essential but adds its own errors, particularly in the presence of noise, to that of the DAC. As stated earlier, this approach compromises the performance of the whole system.

Fig. 10 DAC Located at Distance from Control Loop
The ultimate performance limits of a power converter system can never be better than the digital/analog components used. It is therefore most important that both designers and users appreciate that today the performance limit, in particular for DAC's is at the 16 bit level. The stability of certain key components is not at this moment capable of being improved. Any apparent increase towards higher resolution DAC's will be achieved only by the use of "segmentation" methods. This may give a monotonic performance towards 18 bits but not with a corresponding improvement in linearity. The effects of temperature variations of a few degrees at this level of performance can completely negate the apparent increase in resolution.

7.4 Analog to Digital Converters

In contrast to DAC's, where essentially only one method of conversion is used, there are many methods available for analog to digital conversion. Where high performance is needed (16 bit and above) two principle techniques only remain viable. They are generally known as "successive approximation" and "integrating/multi-slope" methods. Successive approximation uses a high performance DAC in a feedback loop which is controlled by a comparator (Fig. 11). DAC bits are added until the DAC output equals the input voltage to be measured. This method is rapid, typically giving a result in less than 200 microsecs. Unfortunately, any input noise signals will produce errors in the digitised result. A "sample and hold" system is placed at the input of this type of ADC if the voltage to be measured is varying during the digitising process (again adding its errors to the basic ADC). By comparison, the Integrating method provides noise rejection and "signal averaging" automatically. This method, of which there are a number of variants, applies the input voltage to be digitised to an integrator for a fixed period of time, after which an internal reference voltage is applied to ramp the integrator back to its starting point. The time taken to complete this "de-integration" being directly proportional to the input voltage (Fig. 12). The conversion time for this method is very long, being typically several 100 milliseconds for 16 bit accuracy. An additional feature of this method is that by suitable choice of the integration time a very high rejection of mains related frequencies can be obtained (> 40 dB). In current practice, only the multi-slope ADC is able to make precision measurements exceeding 16 bit true accuracy, albeit with the penalty of very long digitising times.

![Fig. 11 Successive Approximation ADC](image)

The problems associated with board layout, grounding etc. and the use of differential amplifiers at the ADC input are exactly the same as mentioned earlier for DAC's. However, the use of a differential amplifier (and probably an input multiplexer) is almost mandatory since within a crate only one "clean ground" point can exist. In this case the input cabling must be kept short and cable screens driven only from their source "low" connection.
8. **PERFORMANCE MONITORING**

In order to maintain confidence in the precision of power converters, constant verification of performance is required. Typically, the output value is read after each output setting, which if implemented in the manner outlined earlier allows a reasonable cross-check of most of the critical components. None the less, regular recalibration should be performed of the DAC, ADC and both output measuring transducers. For power converters which need to have a specified dynamic response, output ripple etc. additional monitoring must be provided. In many cases this will involve some form of signal conditioning circuit, the output of which can be read via a channel of the precision ADC or via an auxiliary ADC. It is preferable to be able to read a digitised value via the remote control system as this can help enormously in diagnosing faulty operation as well as in preventive maintenance. As an example, certain accelerators are very sensitive to short-term output transients caused by power converters. By providing a means to detect and quantify such output transients at distance, a much improved service can be provided.

9. **DIGITAL CONTROL SIGNALS**

Digital control signals modify the internal state of the power converter. Their functions range from ON/OFF control to selection of an ADC multiplexer channel. Any control function which exists the control crate should be buffered and isolated to prevent pick-up and grounding problems. Fail-safe design should be adopted for any control functions which could endanger either the power equipment or personnel. As a general rule it is advisable to limit the rise and fall times of any "power" control signals to prevent unwanted transients. Standardisation of control functions and signals at an early stage can help enormously to produce a rational and consistent user interface.
10. **DIGITAL STATUS**

Digital status signals indicate the existing state of the power converter. In their most simple form they represent the position of contacts e.g. main circuit breaker, thermostats etc. They may be derived from analog values which exceed pre-set limits and so on. A further possibility exists, which has proved to be extremely useful, where they are derived from the dynamic state of the power converter. The majority of digital status signals occur in the "interlock" area and can exceed 100 individual signals. Overall, these signals provide the user with detail within the major "STATES" of the power converter. As many status signals as possible should be provided, since they can speed up operation and maintenance activities enormously.

Again any digital status signals coming from the power equipment must be isolated, screened and preferably filtered.

11. **FAULT DIAGNOSTICS**

Diagnosing faults in power converter systems becomes vastly more important as the amount of material and the physical area of the site increases. The current generation of huge accelerators has highlighted the need to be able to diagnose faults rapidly and effect equally rapid repair. Most of the ability to perform such work efficiently is determined at the system design stage. This has been pointed out repeatedly in the foregoing sections.

With the increased power of modern computer systems it has become feasible to transmit back to the control room as much information as is displayed on the front panels of a power converter. Additionally, important analog signals can be digitised and displayed as waveforms in the control room. This means that a specialist can more often than not determine the cause of the fault or malfunction before leaving the control room or even his office. Ideally, one should be able to pinpoint a defective module or sub-assembly.

The breakdown of faults into the following areas of responsibility can enable the correct specialist to be contacted and much time saved:

i) external interlocks;
ii) mains power and/or auxiliary power;
iii) power circuit and protection;
iv) regulation;
v) remote control and interface;
v) timing signals.

At the hardware level it is advisable to separate functions into individual modules wherever practicable so that exchange of modules affects only a limited functional area. Diagnosis can then be much more easily directed towards module identification, and subsequent exchange.

Clearly, the incorporation of additional hardware to enhance diagnostic capability must in no way decrease either the precision or the basic reliability and must be engineered carefully in these respects.
12. **APPLICATION SOFTWARE**

The software needed to control a power converter system represents an important part of the total programming effort for an accelerator. When the control computers were centralised in the control room, all of the process was performed there. The consequences were that data transfer was considerable and all of the individual particularities of each power converter were taken into account centrally (i.e. via databases). With the advent of decentralised computer systems, much of the specialised treatment necessary migrated to "node" computers situated adjacent to equipment clusters. Today, it is possible to place much of this processing directly in the power converter interface equipment, where it is needed. The use of microprocessors as "intelligent controllers" of power converters has many benefits from the software point of view. The most obvious being that the individual particularities of each power converter can be "hidden" from the higher layers of software. Data transfers and time constraints are considerably reduced, and above all a uniform method of controlling power converters developed.

12.1 **Standardisation and Methodology**

Much pioneering work has been undertaken over the past three years to study means of producing a uniform control of accelerator components, starting with power converters [2]. This preliminary "Methodology" for power converters comprises essential definitions and recommendations which simplify the software to be written but has, in effect, placed the responsibility for most of the work on the power converter specialists (both power and electronics designers). It is therefore important for everyone involved in power converter specification, design and usage to become familiar with this standardisation process and contribute wherever possible.

12.2 **System Evolution**

The ability to improve and adapt applications software over the life of an accelerator is one of the major reasons for using a highly software oriented approach to remote control. While this may be evident for the system as a whole it is equally true at the equipment interface level. Current experience has shown that the ability to reprogram the same hardware to do extra tasks, to correct minor errors or to execute vastly improved algorithms is extremely useful. By a suitable choice of "command language" it is feasible to retain the same high level applications software while making important changes to the interface software. Not only is this approach highly cost effective and quick to implement but it allows the whole system to evolve as naturally as possible.

12.3 **Test and Maintenance Requirements**

Debugging of applications software in today's environment can be extremely difficult if no special provisions are foreseen. The same is true for all aspects of maintenance. It is essential to provide adequate means to control each power converter simply, via a special purpose program developed jointly by the programmers and the power converter specialists. Such a program can incorporate "commands" needed for debugging or maintenance activities which are not available for or needed by the operations staff.

13. **REALISATION**

Much could be said about the realisation of a practical power converter interface. Such work is heavily dependent on the available technology and the experience of the complete design team. The rate of evolution in
electronics, which shows no sign of slowing down, is currently being emulated by spectacular advances in the power field. It should be expected that only the general guidelines which have been outlined earlier will remain valid in a few years time. The major trends however appear to be:

i) a widespread use of Local Area Networks (LAN) for providing the basic communication mechanism assuming that their limited speed can be accommodated;

ii) the increasing use of programmable devices (Microprocessors etc.) embedded directly in the power converters. This may evolve towards direct digital control;

iii) an increase in the overall performance, due to the interaction of digital and analog techniques, as well as improved maintenance possibilities.

A number of recent realisations of power converter remote control systems have been made using some or all of the above concepts. In general, a microprocessor controls the following interface cards via an I/O bus:

i) LAN or communication card;

ii) timing card (for synchronisation);

iii) digital output card;

iv) digital input card(s);

v) DAC card;

vi) ADC card.

The software programming work necessary to make such a system operate in real time is a considerable task and will not be explained further. Such an approach provides considerable flexibility to cope with the individual requirements of each power converter and can provide a most cost effective solution for high performance systems.

However, this interfacing work can only be handled correctly by people who are totally familiar with both the power equipment as well as the remote control, operations and maintenance aspects.

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PROCUREMENT OF POWER CONVERTERS

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ABSTRACT
High performance, modern technology, excellent reliability and low prices are among the key factors of to-day's power converter systems. Procurement of such equipment from industry is not an easy task. Depending on the many boundary conditions, each case will be approached in a different way. On the basis of more than two decades of experience, the problems encountered when acquiring a very large and complex system will be highlighted. Emphasis is put on dealing with problems in a professional and systematic way. The importance of careful preparation will be stressed, starting with a full analysis of the basic accelerator requirements, passing on to the preliminary inquiry, the technical specification and the call for tenders. At the end a few remarks are made on the contract, the manufacture and the testing.

1. INTRODUCTION

Procurement of power converters does not consist only of signing a commercial contract and at a later stage of handing over a bank cheque for payment.

Procurement is the art of conciliating numerous, often contradictory, requirements when acquiring a high-technology power converter system. The many technical facets of such a system have been described during this course. They show clearly that a converter does not only consist of a transformer, diodes and an Ampère-meter. A modern converter system embraces a very wide range of technical fields. The most important are: modern conversion technology, high precision d.c. current measurement, function generators, feed-back theory, interlocks, microprocessor-based performance supervision, real-time operating system and application software. The design team, which masters all these technologies, must also be competent to deal technically and commercially with the power converter industry which manufactures this wide range of products to ensure that they meet the specification. In most cases these firms are selected by competitive tendering. Commercial and general economical aspects play an important role. The constraints of a large accelerator project such as low budget figures, shortage of power converter specialists and a limited time scale transform the procurement of power converters into a real challenge.

Particle accelerators for High Energy Physics are always unique machines operating at the technological limit of all their components. This is particularly true for the power converters. Nevertheless, accelerator users expect high-performance equipment, reliable operation and in addition a considerable potential for future performance improvements.

Experience has shown that the team of power converter specialists responsible for the design and procurement of this equipment will be involved later in operation and improvement. This on-going responsibility
is quite different from standard industrial projects where design, procurement, operation, modifications and repairs are done by teams from different companies, each one making money out of their business. This specific accelerator situation is reflected in the design approach for the equipment as well as in the relationship to industry.

In this chapter I will highlight the problems of procurement of power converter systems for a new accelerator as seen by a CERN engineer. Our experience is based on collaboration with firms from CERN Member States, established in this high-technology field. I will concentrate on technical and managerial problems and will not cover points such as payment conditions, bank guarantees etc. They do not lack importance, but they are outside the framework of this course.

2. **THE OVERALL STRATEGY**

In any accelerator project, the power converters are vital parts closely linked to the other accelerator systems. The converters' performance is dictated on one side by the accelerator requirements and on the other by the needs of the systems they feed with precise d.c. power. Many limits of responsibilities with the other participants in the project have to be defined. Once these points have been settled, the system concept of the power converters can be worked out and a detailed battle plan established, taking into account the numerous, often contradicting boundary conditions.

As an example, let me quote the LEP magnet system. The required performance data are established in close collaboration with the magnet designers and the beam optics specialists. The values of the currents and voltages during filling, acceleration and data taking, as well as the tolerances under static and dynamic conditions, are thoroughly discussed. Quite often an interactive process is required before reaching the final figures.

The limits of responsibilities are established with the various specialists for the buildings, cooling, mains distribution, controls, software, main control room, safety, access control etc.. Precise definitions help in writing clear and complete specifications and avoid expensive additional work during the installation and commissioning phases.

Before working out the system concept and the battle plan, the boundary conditions have to be known. They vary not only from project to project, but also during the project. The most important ones are listed below, in a somewhat arbitrary sequence. The list does not pretend to be exhaustive.

- Competence, experience and size of the team of power converter specialists
- Time scale of projects
- Budget and budget profile
- Technical and financial risks the project management is willing to take
- Overall economic situation in the converter industry
- Interest of industry in participating in the project
- Technical know-how of firms interested in the tenders
- Project extensions, performance improvements
- Safety aspects
- Environmental aspects
- Electrical interference aspects (RFI, mains disturbances)
- Operational budget of the accelerator (manpower, money)
- Operations and maintenance crews.

Keeping in mind the performance to be achieved, the requirements for reliable and efficient operation, the geographical layout of the accelerator and the boundary conditions, the main strategic decisions can then be taken.

In the case of the magnet system these were:
- CERN specialists are responsible for the overall system concept
- They take on the entire performance responsibility
- The design of the equipment takes into account its complete life cycle
- Modern switch-mode technology is used whenever feasible
- Local intelligence based on extensive application of microprocessors is used on all converters
- A comprehensive quality assurance scheme is followed

The main reasons were as follows:

The composition of the LEP power converter team, its knowledge and experience allowed it to take the full responsibility for the design concept for all the converter systems. The team has a long-standing development tradition which allows it to transform new ideas into technically viable projects. This was true, in particular, for the converters of the magnets which cover a power range from 0.7 kW to 7000 kW. The team’s intimate familiarity with all the problems of very high stability in relation to the operation of storage rings made it only natural to take on the responsibility for the vital aspect of performance.

CERN’s continuously declining staff figures made it imperative to design and build equipment keeping in mind the operational requirements. A set of standard electronics for all power converters not only kept the initial cost low but also assured uniformity of documentation and ease of training for the operations and repair crews. Good diagnostics from remote and local control points, operational spares, and good reparability are important design features and were taken into account on an equal footing with excellent reliability. A rigorous, comprehensive quality assurance scheme was followed on all contracts. This included checks of the design, type tests, routine tests, burn-in and complete final testing.

Modern switch-mode technology was introduced on a large scale. Expected advantages were good power conversion efficiency, compact size and modular construction. This results on the one hand in smaller thermal losses to the buildings, hence smaller cooling requirements and on the other hand to smaller buildings. As the power output capabilities of switch-mode or resonance type converters were limited, the higher output powers were covered by mains commutated thyristor equipment.

Every power converter relies heavily on local intelligence, based on microprocessors. They control and supervise each converter in an autonomous way. The incorporated microprocessor transforms the current reference source (in our case a hybrid digital-to-analog converter) into a programmable function generator, handles the incoming and outgoing data stream to the control system and supervises in detail a large number of
internal converter functions. This allows not only the efficient diagnosis of faults from a distance, but also, as a by-product, the analysis of the behaviour of other accelerator systems.

The battle plan was dictated by the very limited financial resources allocated to the converter systems and the short time scale. The split of the overall task into various specifications was guided by technical as well as financial considerations. Each specification should only cover one main technology and therefore a sizeable number of firms should be interested in tendering. In this way, we could obtain attractive prices at an acceptable level of technical risk. Our team of specialists took on not only, as mentioned before, the responsibility of the overall system design, but also the writing of the precise technical specifications, the detailed design of the electronics for the feedback loops, the interlocks, the protection, the local microprocessors and the software.

The battle plan contained also a detailed internal planning. It fixed the dates for termination of the design work, specification writing, review meetings, sending out of tender documents, tender opening, visits to firms, order dates and duration of the various contracts, individual tests in the factory and at CERN, converter commissioning and the start of overall accelerator tests.

Experience has shown that our initial decisions were correct. We did stay within our budget limits; the equipment was ready on time, meets or exceeds the expected performance and runs in a reliable way. It is worth mentioning that the budget figures were established based on contract prices paid in the past. This means that we could beat inflation (in Swiss Francs) over a period of two decades.

3. **RELATIONS WITH INDUSTRY AND UNIVERSITY**

Relations with Industry and University are long-term efforts. In an International Organization one endeavours to cultivate relations throughout all Member States. This gives the opportunity to follow trends in technology and to learn about the capabilities of university departments and industrial companies. It is well known that products based on new technologies are often offered first by small or medium sized firms.

The best way of being informed about industry is of course to be in continuous business contact with it. Conferences, seminars and visits to companies give the opportunity to follow up old and create new channels of information. Ties with universities can be strengthened through development contracts or programmes of "visiting scientists", if such schemes exist in the purchaser's organization.

During the preparation stage of an accelerator project, contacts should be intensified. Special information seminars can be organized and detailed conference papers presented. A successful approach is a buyer's stand at a specialized power converter or power electronics conference or exhibition. A careful presentation with an overall project description, the expected performances and basic block diagrams will help to establish useful relations with the technical staff of companies. A preliminary inquiry is a further step in the right direction. More is said about this subject in one of the next sections.

4. **RULES OF THE GAME**

Profit is the rule of the game in industry. This may be in the form of cash, but also of a gain in know-how, the fact that a company penetrates a new market, gains publicity in participating in a large project and many more.
In order to create a clear situation, the technical specification as well as the commercial conditions have to be clear so that both parties can adhere to them and the contract can be handled in an efficient way.

All rules of the game should be known and discussed fully at the beginning of the project. Some of them are written, others are habits, therefore unwritten. As habits change, they undergo the same process. The written rules, such as financial rules, general rules of purchase etc. have to be studied carefully in order to fill in certain gaps or give emphasis to some particular problems. To find out all about the unwritten rules is certainly a demanding job. A course on "How to treat your boss" may give some useful hints.

The legal aspects are important too. You certainly have heard about the patent laws, national laws, labour laws, Roman law and English law, all of them taught at University level. But in everyday life, the "law of the jungle" has its importance. When deciding about payment conditions and schedules this point should not be forgotten.

5. CALL FOR TENDERS

5.1 Preliminary phase

Any important call for tenders will go through a preparation phase. This may take the form of a preliminary inquiry which informs industry about future requirements and at the same time serves as a market review. The main document of the preliminary inquiry is a short technical specification of the system or part of it, preceded by a concise overall view of the complete accelerator project. It is wise to split up the preliminary inquiry in the same way as one will do with the final call for tenders. A request for the firms' documentation and a detailed questionnaire allows technical and commercial information about the firms to be collected.

The preliminary inquiry may also be used for selecting companies. In this case the selection criteria have to be spelt out very clearly. They may include financial aspects such as turn-over in the field of converters as well as technical aspects, such as technical expertise in a specific field. It is important to mention that only firms replying to the preliminary inquiry and meeting the selection limits will be invited to the final tendering. The initial large lists will therefore be reduced to a "short list". Such a "short list" of firms has the advantage of inviting only manufacturers with expert knowledge. The selected tenderers from their side see their chances for success increased. At a later stage, the purchaser may request, before placing the contract, proof of the technical expertise of the tenderer making the lowest offer. This will take the form of a qualification prototype meeting the specification.

A follow-up of the preliminary inquiry may consist of visits to some of the firms which replied. Discussions with specialists will then be based on a written document which, although condensed, considerably increases the success of the visit. At this stage it is important to sound out design, development, production and quality control facilities. They are key factors for obtaining low price, good product quality and reliable delivery.

5.2 Tender documents

Numerous technical and commercial papers as well as various annexes form the file of tender documents. The technical specification is the most important one followed by the questionnaire. These documents have a
wide variety of functions in the course of their useful lives. Therefore a later section deals in detail with the technical specification and the technical questionnaire.

On the commercial side, the total quoted price is of course the most important financial information supplied by the tenderer. The price breakdown, which is always requested, should be such that its information allows the technical content of the offer to be cross-checked. Prices for the options and their validity have to be included. Tenderers' comments on the delivery schedule should be encouraged.

The tender documents must remind the tenderers about the selection criteria applicable and may indicate the tender opening procedure. To inform the tenderers about the internal administrative procedure after tender opening is a useful step.

A reasonable period of time should be given to the firms for establishing a technically sound offer. Direct information to the technical people may reduce the length of time documents lie on administrators' desks.

6. **TECHNICAL SPECIFICATION**

6.1 **Preparation of the specification**

Before writing a technical specification, it is worthwhile carefully documenting the basic information supplied by other groups of the project, as well as the design decisions and computations. This documentation should include a short description of the history, a discussion of the possible solutions and their pros and cons. Detailed design calculations can be added, the tolerances of the various components given and values for environmental conditions and reliability considerations written down.

A precise time schedule will be a "must" for the follow-up of the various stages of writing the drafts and the final versions of the technical specification, its annexes, questionnaires and the various additional documents.

6.2 **The specification**

The technical specification is used by many people and has to fulfil a wide variety of functions over its useful life. The most important are listed below:

- Firstly, it must provide quick and precise information to a potential supplier on the level of "Decision Makers". It must give technical content, quantities and time scale.
- Secondly, it serves as the basis of the tenderer's detailed technical offer and the "price tag". All the information needed for this purpose must be clearly spelt out.
- Thirdly, it is the main working document for the duration of the contract. It serves the design office, the production department and the subcontractors as well as the test and installation teams. Last but not least, it helps the engineer responsible for the contract, to get the product and services his organization ordered and will pay for!
The technical specification is a contractual paper and therefore has to be written with this aspect in mind. It will often be read on its own, separated from the other parts of the tender documents. It must be a complete document in its own right and therefore will overlap with some of the other tender documents.

The structure of the specification reflects the multiple use of this document. It will therefore contain at least the following main chapters or annexes:
- Short summary (technical data, quantities, delivery schedule)
- Introduction to the accelerator project
- Description of the general environment (buildings, cooling, etc.)
- The hard core of the detailed technical specification of the product and services the tenderer has to supply.
- A technical summary list of all the numerical values of the equipment with their references to the chapters
- Quality assurance scheme to be applied
- Detailed testing procedure, type tests, routine tests including burn-in, definition of what is done in the factory of the contractor or subcontractors or at a special test facility
- Quality requirements for certain materials (cables, etc.)
- List of items delivered to the contractor for questions of standardization
- List of items to be purchased from a given supplier
- Documentation required
- Nomination of contract engineer
- Options: quantities, rate of delivery, validity
- Transport to and erection on the accelerator site
- Facilities on the accelerator site
- Testing on accelerator site on dummy and real load
- Guarantee work, rapidity of repair service
- Safety rules applicable
- Access conditions to the site, working hours, restrictions due to overall planning.

The "hard core", i.e. the real subject of the specification, must be treated in great detail. Block diagrams, circuits and numerical values, including tolerances, have to be given in a language familiar to specialists. Neither a design engineer for a power converter, nor a test engineer, will understand notions such as width of stop bands, betatron oscillations and the like. He must have information, figures and tolerances which are familiar to him and which he can achieve. Very often these are reached thanks to a special effort made first at the design then at the manufacture and finally at the testing stage. Accelerator requirements are often harder to meet than ordinary industrial ones.

A good drawing is more informative than a long description. Layout drawings with dimensions, figures with waveforms for currents and voltages will pass the message more easily. The specification will be read and used by people of different educational background and mother tongue than those of its author. Maybe future specifications will be in the form of "comic strips". Their chance of being studied in detail by everyone involved in the contract will increase considerably.
The annexes must be chosen carefully. They must include international recommendations (such as IEC standards), national and internal rules and prescriptions. They also cover safety aspects, measurement technology, environmental aspects, radiation protection, site access rules, local conditions for work permits, hotel accommodation etc.

A power converter contract will be handled by many departments of a firm. From outside it is difficult to know all the people and to follow the contract in an efficient manner. In this case a "contract engineer" is an essential person. It is vital to nominate such a person! He must have easy access to all the important people of the various departments and be in the factory most of the time. He will make sure that the client is continuously informed and that his requests will get through to the people concerned.

Flexibility, restructuring and subcontracting are part of the everyday vocabulary of modern industrialists. The question has to be asked where the equipment will be made and tested, with the full address of the manufacturing department. The same is true for the subcontractors. It is compulsory that manufacturing facilities as well as subcontractors can only be changed with the purchaser's written agreement.

Delivery schedules should be realistic for both sides. Short delivery times increase the price. A delivery schedule which is too short will leave the purchaser with adventurous firms or no tenderer at all - two possibilities to be avoided. An early start will increase the chance of getting the product on time at a reasonable price.

Options are very useful and often a "must". They make sure that additional spares or extensions of the project will have the same material for a known price. The duration of validity of the option clause as well as the rate of delivery must be indicated and should be in line with the main contract.

The technical questionnaire is also an important document. A great deal of attention should be devoted to its formulation. Questions should be formed in such a way that only a reasonable design study can answer them. The questionnaire is the first step in checking whether the offer meets the specification. It also allows comparison of offers from different tenderers. Requests for data-sheets, sketches etc. can be included as well as manufacturers of the components purchased, their type numbers etc. During the execution of the work, the questionnaire is an additional and useful tool for the contract supervision.

Guarantee clauses are defined in the general purchase conditions. They are formulated in such a way that they satisfy many different cases. For a specific application more details may be needed which have to be added to the specification. Repairs under guarantee may be such a case. How rapid must the intervention be? If a 24-hour repair service is requested, this has to be specified. Otherwise the manufacturer may have no stock of spare parts and will first have to manufacture or order the replacements for the broken components.

Most accelerator departments have a permanent operation/maintenance crew, which is responsible for the efficient running of the "machine", including the converters. This unloads the manufacturer considerably from his guarantee duties. Therefore, only equipment with systematic faults will normally be returned to the factory. The reduced extent of the guarantee may well be reflected in a substantial negotiated price reduction - an idea worth discussion.
Last but not least, the educational aspect of the specification writing is important to the design crew. By formulating, reading through and discussing the draft and the final specification, all members of the team go through a very useful learning process which will reflect itself in many improvements. The specification must be read as through the eyes of the tenderer. How will he interpret it? What is his solution for meeting the specification? May he offer, by strictly adhering to the text, a solution the user dislikes? In certain cases, less prescription may do more good!

7. THE CONTRACT AND ITS FOLLOW-UP

7.1 Contract preparation

Tender opening is always an exciting moment full of surprises. After careful analysis of the tenders by the design team, possibly assisted by people from outside, visits to the firms will follow. If time allows, the two or three most interesting firms should be visited, including their main subcontractors, in order to discuss the tender in detail. Thus the customer makes sure that the specification is fully understood and the offer is complete. The client will meet the company's contract engineer, the contact man for several months to come! References are very useful and should be checked thoroughly.

A further stage in the selection process, as mentioned earlier, ahead of the contract signature, may be the construction and approval of qualification prototypes. This will show whether the lowest tenderers are in a position to manufacture the goods to specification. Although this process may at first look time-consuming and expensive, in the end it often saves money and trouble.

The internal approval mechanism of the client's organization will choose the firms which satisfy all the selection criteria, taking into account the prices, offers which meet the technical specification, and the delivery times. The company's experience has been proven, if possible, by a successful qualification prototype.

The patent situation is, in general, covered in the conditions of purchase. There, the emphasis is put on the protection of new ideas which will arise during the execution of the contract. A point which is worth bringing up is the coverage by existing patents of the products the tenderer is proposing to use during the contract execution.

7.2 Manufacture

Regular contacts with the firm will be established from the very beginning. They will assure that the company's design follows closely the specification. They also make sure that difficulties, which may arise during the contract period, can be immediately spotted and corrective action taken. The main features of the specification have to be repeated again and again. If new people join the contract work, one has to be aware that not all of them will know all the details of the specification.

In the present industrial context, main contractors will purchase major items from subcontractors. Transfer of the information contained in the specification to subcontractors is of vital importance and it is worth checking carefully this aspect.
7.3 Testing

Quality assurance during the contract is a continuous task. All inspections, burn-ins, type and routine tests have to be carried out and supervised conscientiously and thoroughly as stated in the specification. Even for large series, defects may creep in during production. Only complete inspections and tests of all the equipment assure good quality. Wherever possible, factory tests should reflect exactly the service conditions on site. Repairs or up-grades during operation are much more expensive than thorough factory tests. A little extra time spent in testing can save much additional cost and embarrassment later.

Despite all these precautions, early failures will appear during the first hundred hours of operation. Whenever possible, accumulate operating hours well ahead of the beginning of the accelerator commissioning.

8. Conclusions

Procurement of power converter systems requires expert knowledge in a wide variety of fields and special care to be paid to detail. Nevertheless, problems do not always arise where expected. Contracts which appear difficult at the beginning may run very smoothly whereas others, seemingly straightforward, lead to unexpected trouble.

And now good luck for your next procurement of power converters; an exciting challenge is ahead of you!
RESONANT EXCITATION OF SYNCHROTRON MAGNETS

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Abstract
Principal and extended networks for resonance excitation, their optimal design and their desirable and undesirable characteristics are described. The generation of combined waveforms is explained and finally the basic properties of power sources, tuning and protection devices and a selection of automatic control systems are presented.
This is a summary and review of the methods which are known so far.

1 Introduction

Excitation of synchrotrons having higher repetition frequencies $\omega$ for their magnet cycle runs into difficulties if a normal linear ramping mode is used because the electric power for energizing and deenergizing the magnets increases with $\omega$ and with it the load of the grid and the distortion of the voltage. This implies that a method of storing the energy of the magnets, which is needed only pulsewise during each period and which is able to work at frequencies $\frac{2\pi}{T} \geq$ one or few Hz, has to be introduced. Therefore oscillatory electric means are required, which use electric storage elements, such as, at least capacitors. In case of resonance, only the AC power remains to make up for the losses of the system. The load of the mains is therefore reduced drastically but is one phase AC-wise. Depending on the sensitivity of the other consumers connected to the mains, the active power swing has to be decreased by energy storage elements inserted in the power source or by compensation methods. The minimum useful repetition magnet frequency for resonance excitation depends on:

- economic considerations, i.e. costs of the resonance excitation network + power sources against costs for a converter-inverter device for ramp excitation + dynamic reactive power compensation for the uncompensated active power swing;
- influence of the active power swing on the phase angle of the mains voltage $\Delta \varphi$ and permissible value of $\Delta \varphi$ in case of ramp excitation. The compensation of this phase angle deviation can be very expensive in comparison to the main installation costs, especially in the case of smaller plants.

The superiority of the ramp excitation, namely generation of different curve shapes, is partly compensated for by the possibility to produce waveforms from a combination of portions of harmonic shape with portions of a flat character, of waveforms of a superposition of excitation modes of two frequencies, or waveforms resulting from oscillatory modes with two different frequencies.
2 Fundamental Mode Excitation

2.1 Layout and Properties

For acceleration only an almost unidirectional magnet current shape is necessary and therefore a DC bias current $I$ is added to the sine-waveform, getting now the function of the magnet excitation current $i_m$,

$$i_m = I(1 - \varepsilon \cos \omega t)$$  \hspace{1cm} (1)

The value of $\varepsilon$ depends on the injection field required. If the latter is very low, $\varepsilon$ is chosen to be a little more than unity. Introducing the term for maximum and minimum magnetic field $B_{\text{max}}$ resp. $B_{\text{min}}$, the magnet field excitation shape $B$ is given by

$$B = \frac{B_{\text{max}} + B_{\text{min}}}{2} - \frac{B_{\text{max}} - B_{\text{min}}}{2} \cdot \cos \omega t$$  \hspace{1cm} (2)

Figure 1 shows the basic circuit for such an excitation function, here being comprised of two power sources for making up the DC and AC losses in the system. Thus the mains are only charged by a small part of the alternating power given by the quality $Q$ of the resonant circuit. For example, for 50 Hz excitation frequency, $Q \geq 50 \ldots 100$ for magnet circuits of accelerators with particle energies of more than 1 GeV. The mode of AC supply -- either parallel or series -- determines the resonant character of the same kind. Series excitation requires the use of transformers, series-connected to $C_1$ resp. $C_2$. It must be pointed out that the choke $L_1$ is necessary to bypass the DC and thus the capacitor $C_2$ resonating with $L_1$ AC-wise. The costs for these additional elements are normally no greater than those of a pure AC excitation since in this case the AC voltage is doubled and the size of the capacitor $C_1$ has to be quadrupled. Apart from this, there are more important objections to such a mode of excitation such as a much higher $\frac{dB}{dt}$ at injection and a wasted half-period of the magnet cycle.

As shown in the figure, all magnets are connected in series to achieve equal magnet current everywhere. With larger synchrotrons, especially for the dipole magnets, a simple series connection of all magnets would result in a too high voltage to ground since, in the given relation winding current $I \times$ number of turns $n$, not any ratio of $I$ and $n$ can be chosen. To meet this difficulty, the circuit elements are connected in the so-called White Circuit, of Fig. 2 [1,2]. Here the number of magnets allowed -- with regard to the voltage -- is mutually connected in series with their resonant capacity $C_{11}$, while a parallel resonant circuit, consisting of one DC bypass choke $L_{27}$ and the capacity $C_{27}$, which belongs to it, is connected in parallel to $C_{11}$.

These four elements form one group of the White Circuit, the number of groups depending on the quantity of magnets. One group is split for the connection of the DC source which is center-grounded and thus the midpoint of each magnet group -- and therefore capacitor group -- is virtually grounded under steady state conditions.

The function of the circuit just described with regard to the production of the desired magnet current
\(i_m\) is independent of the size of \(L_1\), resp. \(L_2\) or the ratio \(\alpha = \frac{I_{L1}}{I_{L2}}\) resp. \(\frac{I_{L2}}{I_{L1}}\). Since the DC determines the direction, the choke current yields \(i_{ch} = I_1 (1 + \alpha \cdot \cos \omega t)\) and the stored energies \(W_m, W_{ch}\) and \(W_c\) for the magnets, choke and capacitors respectively, are, with \(\varepsilon = 1\),

\[
W_m = \dot{W}_m \left(\frac{3}{8} - \frac{1}{2} \cos \omega t + \frac{1}{8} \cos 2 \omega t\right)
\]

(3)

\[
\dot{W}_m = 2 L_1 I_1^2
\]

(4)

\[
\frac{W_{ch}}{W_m} = \left(\frac{1}{4 \alpha} + \frac{\alpha}{8} + \frac{1}{2} \cos \omega t + \frac{\alpha}{8} \cos 2 \omega t\right)
\]

(5)

\[
\frac{W_c}{W_m} = \frac{1 + \alpha}{8} (1 - \cos 2\omega t).
\]

(6)

The sum of the equations (3), (5) and (6) is a constant as is required. Figure 3 shows the waveforms of the currents, the voltage and the stored energies for \(\varepsilon = 1\) and \(\alpha = 0.5\).

The determination of \(\alpha\), i.e. the size of the choke, results from economic considerations. For this, we have to consider the total costs for the choke \(L_1\) and the capacitor \(C_2\) as follows:

**Choke investment costs** \(C_{ch}\), using the price per kVA for 50 kHz equivalent choke sizes \(c_{ch}\)

\[
C_{ch} = c_{ch} \cdot \frac{1}{2} \omega_{50} L_1 I_1^2 \cdot \frac{1}{\alpha} (1 + \alpha \varepsilon)^2 \cdot f_c
\]

(7)

where \(f_c\) is the correction factor for current waveforms where the maximum value \(I_{ch} \neq \sqrt{2}\) rms-value \(I_{ch}\)

\[
f_c = \sqrt{2} \cdot \frac{I_{ch}}{I_1} = \sqrt{2} \frac{\sqrt{1 - \frac{\alpha^2 \varepsilon^2}{2}}}{1 - \alpha \varepsilon}.
\]

(8)

Thus we get:

\[
\frac{C_{ch}}{c_{ch} \omega_{50} L_1 I_1^2} = \frac{1}{\sqrt{2}} (1 + \alpha \varepsilon) \sqrt{\frac{1}{\alpha^2} + \frac{\varepsilon^2}{2}}.
\]

(9)

**Choke energy costs** \(C_{ch}\): using the price per kWh losses \(c_p\) derived from the price per kWh and the lifetime, e.g. 30,000 hours:

\[
C_{ch} = I_{ch}^2 \cdot r_{ch} \cdot c_p
\]

(10)

where \(I_{ch}\): choke current rms-value and \(r_{ch}\): equivalent choke resistance representing all losses. Using the
quality of the choke \( Q_{ch} \) at a line frequency of 50 Hz \( = \frac{\omega_{op}}{2\pi} \) and \( Q_{ch} = \frac{\omega_{op} L_2}{r_{ch}} \) we get

\[
\frac{C_{chp}}{c_p \omega_{op} L_1 I_2^2} = 1 + \frac{1}{Q_{ch}} \left( \frac{1}{\alpha} + \frac{\epsilon^2}{2} \right).
\]  \( \text{(11)} \)

Capacitors \( C_2 \) investment costs \( C_{ci} \) and the specific costs \( c_s \) per kVA using a similar equivalent 50 Hz size, derived from the real costs at cycle frequency \( \omega \) multiplied by the factor \( \frac{\omega}{\omega_{op}} \).

\[
C_{ci} = c_s \cdot \omega_{op} L \epsilon^2 \cdot \frac{\epsilon^2 \cdot \alpha}{2}
\]  \( \text{(12)} \)

Capacitors \( C_2 \) energy costs \( C_{ce} \): Since the losses of power capacitors are given in W per kVA for a determined frequency, energy costs may be accounted for by an extra charge to \( c_s \).

The sum of the related costs \( C_r \) of the bypass-parallel resonant circuit \( L_2, C_2 \) is therefore:

\[
C_r = \sum \frac{C_i}{C_e \omega_{op} L_1 I_2^2} = \beta + \frac{1 + \alpha \epsilon}{\sqrt{2}} \sqrt{\frac{1}{\alpha^2 + \frac{\epsilon^2}{2}}} + \alpha \frac{\epsilon^2}{2} \left( 1 + \frac{\gamma}{Q_{ch}} \right) + \frac{1}{\alpha} \frac{\gamma}{Q_{ch}}
\]  \( \text{(13)} \)

where \( \gamma = \frac{\omega_{op}}{\omega} \) is a constant value since the specific costs per kVA, including losses, are constant for a determined frequency and voltage. Only the two parameters \( \beta = \frac{c_{ch}}{c_s} \) and \( Q_{ch} \) then remain to be calculated.

As example the optimum of \( \alpha \) is to be found out for a 50 Hz repetition rate, \( \epsilon = 1 \), and the parameters \( \beta = 0.4, 1.0, 2.5 \) and \( Q_{ch} = 200,600 \).

The electric losses of 50 Hz all-film power capacitors without discharge resistors are 0.25 W per kVA. Assuming 30,000 hours of total operation time and 0.15 DM/kWh average power costs, the extra charge to the specific capacitor price due to capacitor losses is 1.125 DM/kVA. Therefore we use \( c_s = 10 \) DM/kVA as the specific capacitor cost and so obtain \( \gamma = 450 \).

Figure 4 shows the related costs \( C_r \) as a function of \( \alpha \) with the two parameters \( \beta \) and \( Q_{CH} \). Evidently the cost minimum lies between \( \alpha = 1 \) and 1.2 over the whole range of parameters.

Investment costs \( C_{chp} \) of the choke only have a minimum for \( \alpha = \sqrt{2} = 1.26 \), and the maximum of the stored energies of both the elements \( L_2 \) and \( C_2 \) taken as a basis yields an optimum of \( \alpha = \frac{1}{\sqrt{2}} = 0.707 \).

Even for early applications of the White Circuit, the idea arose to combine the \( n \) single reactors of an \( n \)-piece circuit to one unit with \( n \) separated windings, i.e. the magnetic circuit is common. The reason for this is based on the rule of the volume \( V \) (or weight) of electric machines or transformers as a function of their power \( P \) for equal densities of current and magnetic flux:

\[
\frac{V_1}{V_2} = \left( \frac{P_1}{P_2} \right)^{\frac{1}{n}}.
\]  \( \text{(14)} \)

The same function is valid for the power losses. Thus for \( n = 10 \) the expense is reduced to 56 %. Such joint reactors have been used in the form of:
a) a ring choke [3]

b) an air core choke with an outside shielding of aluminium to screen the alternative flux [4]

c) an air core choke like b) but with an iron yoke as the return path for the magnetic flux [5]

In regard to design a): Figure 5 shows a ring core choke in a polygonal assembly of 12 trapezoidal laminated iron cores bearing the 12 coils as installed for the Cambridge Electron Accelerator (CEA). This arrangement is symmetrical. Not shown in this principal representation is the outer clamping device. Improvements in design would have been necessary to place the split coils of each sector on both sides of the gaps to avoid magnetic stray fluxes. But that makes the whole assembly rather complicated.

Figure 6 demonstrates the design principle c of the joint inductor used for the DESY synchrotron (a similar choke was also later used for NINA [6]). It is based on the idea that in a cylindrical stack of n coils and \( n - 1 \) distances \( d \) between them, symmetry conditions are achieved for an arrangement with the distances \( \frac{d}{2} \) between the end coils and the iron yoke covering their whole diameter (magnetic mirror). In practice this is not realistic since more space than \( \frac{d}{2} \) is used to house the clamping device between the iron mantle and the end coils. Therefore the number of turns has to be enlarged from the middle to the ends of the coil stack.

With the symbols of Fig. 6, and those which have already been used, in addition current density \( J \) and winding filling factor \( f_w \), the maximum energy stored of such a choke of window-frame type is:

\[
W_{ch} = 2 \mu_0 \pi \left( \frac{J f_w}{f_r} \right)^2 \delta r_a \left[ \frac{1}{6} - \frac{2}{3} \left( \frac{r_i}{r_a} \right)^3 + \frac{1}{2} \left( \frac{r_i}{r_a} \right)^4 \right]
\]  

(15)

If \( \frac{r_i}{r_a} \) is sufficiently small, the term in parentheses is a constant and the formula is simplified. So the two parts of the choke, weight of copper and weight of iron yoke, can be expressed as a function of only the two variables \( \delta \) and \( r_a \). Taking into account the capacitors, which depend on \( \delta \) only, the minimum of the overall costs of the bypass-parallel resonant circuit can be calculated from an equation with two variables in a similar manner to that described above.

### 2.2 Undesirable Characteristics of the White Circuit

#### 2.2.1 Earth Faults and Stray Capacitance

The desirable characteristic of the White Circuit, i.e. the voltage distribution around the ring with excitation in the fundamental mode, is shown in Fig. 7 for the undisturbed state. Under earth-fault conditions the voltage distortion depends on the grounding resistance and the fault resistance to earth. In the worst case, twice the normal voltage to ground occurs. The influence of stray capacitance can be eliminated or reduced for the fundamental mode in a similar way to the shielding methods used in electronic experiments:

- capacitor banks: the capacities from coating to box are connected in parallel to the main capacitance by connecting the boxes to the real or artificial centre of the banks. The boxes have to be insulated
from ground. Then the capacities from boxes to earth become efficient, but with the boxes virtually earthed in normal state for the fundamental mode the current flowing to ground is low.

- cables: the cables must have an insulated screen which has to be connected to the capacitors if this is allowed according to the safety rules. In case only strongly grounded cable screens are permitted, tuning is necessary to obtain a minimum leakage current of each group to ground, see circuit diagram of Fig. 7.

- chokes: their primary to secondary winding capacitance allows leakage currents to flow parallel to the magnet groups due to the paralleled primaries. This can be avoided if opposite parts of different coils have the same voltage, i.e. there is no potential difference between them. Another method could be double screening between primary and secondary windings. The screen near the secondary has to be connected to the mid-point of the adjacent capacitor bank, whereas the other one is grounded.

Remaining capacitive leakage currents flowing to the real or grounding point cause asymmetries in the magnet currents.

2.2.2 Spurious Modes

The White Circuit is able to resonate in spurious modes if symmetry is not ensured. Symmetry conditions are fulfilled if:

a) the mutual inductances between the bypass chokes are equal (including zero), and

b) the capacitances and the magnet inductances per group are equal, and

c) the power source is coupled symmetrically to each group.

It has to be stated that a symmetric windings arrangement in the case of a combined reactor, as mentioned above, does not also mean symmetry in mutual coupling. At the bottom of Fig. 7 the voltage distribution in a three-cell White Circuit (10 : 1 scale) containing a joint inductor with asymmetrical coupling is shown [6]. Figure 8 shows diagrams of frequency responses of a simulated three-mesh White Circuit with weakly coupled \( k < 1 \) parallel current sources and a fundamental resonance frequency of nearly 50 Hz.

If the coupling coefficient according to a) or c) is unity or very near to unity, no other symmetry conditions have to be observed to avoid spurious modes. This can be attained by distributing each coil over the whole length of windings of the joint reactor, see Refs. [6,7], or by using closely coupled and parallel connected auxiliary windings, so-called equalizer windings. The latter method is the most practical one and was used from the beginning. In the case of independent bypass inductors \( k = 0 \), closely coupled primary windings, which have to be in parallel, are also necessary for each of them. The parallel-connected primaries can also be used to supply the AC-power, if a parallel resonant characteristic of the whole circuit is desired.
2.2.3 Delay Line Modes

In the case of higher frequencies, the White Circuit acts as a ladder network consisting of the magnet inductances in the line shunted by the stray capacitances to earth. Since these are much smaller than the main capacitances, the capacitor banks may be replaced by short circuits. Thus the whole circuit is able to resonate in delay line modes. In the loss-free case the resonances take the form of standing waves with voltage nodes at the grounding point. The frequency of the fundamental is then given by:

\[
 f_0 = \frac{\omega}{2\pi} = \frac{1}{2\sqrt{LC}}
\]  

(16)

assuming symmetrical distribution of the total stray capacitance C (L: total magnet inductance). Excitation of delay line modes is mainly achieved by using additional pulse power sources, e.g. for beam bump systems. However, the main power for the supply of the fundamental AC component is delivered at the same time to all groups of the network and, in this case, delay line modes can only be excited by it if the ladder network is not uniform [8]. A successful means of eliminating or reducing the effects of delay line mode resonances is the installation of damping loops [8]. As shown in Fig. 9 they consist of series-connected magnet backleg windings closed by a resistor. The series connection must have reverse polarity in order to cancel the voltages induced by the main power supply. The voltages due to the fundamental or any odd harmonic of the delay line resonance are in phase. Maximum damping is achieved by a resistance equal to the reactance of the two magnet windings at the frequency to be damped.

3 Compound Waveforms

So far, two applications are known where a simple harmonic (plus bias) shape of the magnetic field is not sufficient:

- The requirement of a constant or nearly constant particle energy during a given time, mostly at the maximum or the injection field makes it necessary to produce a waveform having a flat-top or flat-bottom part or both.

- The effort to save rf power leads to a shape having a slower rise and a faster decay with the same periodic time as the simple waveform [9].

There are two methods available to produce combined waveforms: superposition of harmonics and switching of energy storage elements.

3.1 Superposition of Harmonics

Figure 10 shows the effect of superposed harmonic components on the shape of the magnet current, namely a \( \frac{\sqrt{2}}{8} \) portion second harmonic for rise time extension, and an 8.5% portion fourth harmonic for forming
flat tops (both related to the AC fundamental). The extension of the existing circuit by adding two independent energy storage elements allows it to ring at three resonance frequencies.

Representing one of many possible solutions, Fig. 11 shows a proposed network [10] which figured as one cell of the former DESY I White Circuit with series supply for the AC fundamental. Here, and for all similar networks, tuning of resonance circuits to avoid interactions between sources of different frequencies arises as an additional difficulty. Therefore a combination without internal resonance circuits must be preferred. This can be achieved by using dual reactors and applying a compensation method using the primaries of the existing chokes and one winding of the new chokes for separating the sources of different frequencies [11]. Simplified circuit diagrams for a one-cell configuration and without DC component are shown in the Fig. 12 containing two methods, namely: dual chokes in series with the magnets – Circuit A – or in series with the capacitor banks – Circuit B – . Setting \( y = \frac{L_{21}}{L_{11}} \) and \( n \) as ratio of transformation from primary to secondary of the main inductor and assuming unity coefficient of coupling for both main and additional reactor (\( M_{21} = \sqrt{L_{21} \cdot L_{21}'}; \ M_{31} = \sqrt{L_{31} \cdot L_{31}'} \)), we get the terms of Table 1.

<table>
<thead>
<tr>
<th>Circuit</th>
<th>A</th>
<th>B</th>
</tr>
</thead>
<tbody>
<tr>
<td>Condition for compensation</td>
<td>( \frac{M_{21}}{L_{11}} \cdot n \cdot (1 + y) )</td>
<td>( n \cdot \frac{1}{1 + \alpha} )</td>
</tr>
<tr>
<td>Primary inductance of add. choke</td>
<td>( \frac{L_{21}}{L_{11}} \cdot n^2 \cdot \left( 1 + \frac{1}{y} \right)^2 )</td>
<td>( n^2 \cdot \frac{1}{y(1 + \alpha)^2} )</td>
</tr>
<tr>
<td>Harmonics input-current</td>
<td>( \frac{i_{2n}}{i_{1n}} )</td>
<td>( \frac{1 + \alpha}{n} )</td>
</tr>
<tr>
<td>Harmonics input-impedance</td>
<td>( \frac{\omega L_{11}}{L_{11}} \cdot n^2 \left( 1 + \frac{1}{y} \right) )</td>
<td>( j \omega L_{11} \cdot n^2 \left( \frac{1}{y(1 + \alpha)^2} + \frac{1}{1 - \alpha} \right) )</td>
</tr>
</tbody>
</table>

The size of the additional elements for superposition is a matter of optimization. Using \( \delta = \frac{i_{pk}}{i_{m}} \) – the ratio of the peak value of the superimposed harmonic magnet current to the peak value of the total magnet current – as the coefficient of superposition, we get the following terms for maximum stored energy in the case of circuit A:

**Additional Chokes**

\[
\hat{W}_c = \frac{1}{2} L_{21} \hat{i}_{2m}^2 + \frac{1}{2} L_{21}' \hat{i}_{2n}^2 + M_{21} \hat{i}_m \hat{i}_n \\
\frac{\hat{W}_c}{\frac{1}{2} L_{11} \hat{i}_{1m}^2} = y + \left( 1 + \frac{1}{y} \right)^2 \delta^2 + (1 + y) 2\delta
\]

**Additional Capacitors**

\[
\hat{W}_{c2} = \frac{1}{2} L_i \hat{i}_{2n}^2
\]

where

\[
L_i = L_{11} \cdot n^2 \left( 1 + \frac{1}{y} \right)
\]

\[
\frac{\hat{W}_{c2}}{\frac{1}{2} L_{11} \hat{i}_{1m}^2} = \delta^2 \left( 1 + \frac{1}{y} \right)
\]
From these two terms as a function of \( y \), an optimal value of \( y \) can be found in a similar way to that described in section 2.1. For the transfer into kVA-units, the \( f_\gamma \)-factor of the additional choke may be near \( \sqrt{2} \) by the influence of the primary winding which carries a sinusoidal current. — For this optimization the influence of the superposition on the main capacitor bank \( C_{11} + C_{21} \) also has to be considered. Here the stored energy of the fundamental mode can be increased by the influence of the additional choke, but there is no general rule for that.

3.2 Wave Shape Modification by Switch Modes

Proposals for forming flat-top periods by short-circuiting the capacitor banks if their voltage is zero, i.e. the magnet current is at its maximum, have been known for a longer time [12]. In this case, the White Circuit was modified by the insertion of inductors in series to the capacitor banks [13]. This made it possible to make a short-time flat-top symmetrical to the original wave shape, but in practice operation in this mode caused transient voltages of higher frequencies [14]. The other reason for using such a system is the possibility of self-commutation at the end of the flat-top time for thyristor switches. Figure 13a shows the simplified circuit diagram for producing short-time flat-tops. Thyristor \( Th_1 \) makes the short which produces the flat-top. During this time, \( C \) is discharged by damped oscillation via \( L \) and \( Th_1 \). When \( C \) has its maximum negative voltage thyristor \( Th_2 \) is fired and an inverse current starts to interrupt the \( Th_1 \) current. However, the remaining charge of \( C \) must be sufficiently high and therefore this method can only be used for short-time flat-tops. Those of long duration require interruption of the capacitor discharge current and therefore more semiconductor elements. The latter arrangement is shown in Fig. 13b and its operation in Fig. 14, assuming that \( \alpha = 1 \) and \( \varepsilon = 1 \), i.e. \( L_{11} = L_{21} \). This layout results in the current \( i_{ch} \) being zero during the flat-top. For other values of \( \alpha \), \( i_{ch} \neq 0 \) at the start of the flat-top and then drops due to the circuit resistance. On the other hand, the magnet current is stable during flat-top time and this would result in different initial conditions when the regeneration phase starts. During this phase, energy supply is necessary, e.g., by a current pulse. It should also be noted that flat-top operation without the additional inductors is possible but, while the main circuit would be simplified very much, additional equipment for forced commutation of the thyristor switches is needed.

In contrast to the switch-on mode forming a flat phase, switch-off operation is applied to obtain an increase in the rise time. As proposed in Ref. [9], fractions of the capacitor banks are separated when they have zero voltage or the magnet current is at its maximum. Interruption has to be achieved by forced commutation of the thyristor switches or by a combination of gate turn-off thyristor – GTO – and thyristor [15] as shown in Fig. 15 showing currents and voltages for a 1:3 frequency ratio for which \( \frac{8}{9} \) of the total capacitance has to be switched off. The remaining \( \frac{1}{9} \) is stressed to the threefold voltage. Therefore it has to store the whole reactive energy and the capacitor banks have to be designed for \( \frac{12}{9} \) in terms of energy stored related to a single frequency excitation. As shown in the diagram, the operation mode of the capacitors is pulsewise with different pulse lengths and unidirectional voltage pulses and this
influences their layout. This arrangement can be more favourable in comparison to pure AC capacitors. The semiconductor arrangement shown protects the GTO from the forward off-state voltage while the diode limits the reverse voltage.

The combination of the two switch modes, i.e., one switch in series to a fraction of the capacitors and one parallel to them was first described in Ref. [16] to produce an additional flat-bottom phase for injection. To overcome the difficulty of decreasing choke current (see above), the insertion of one power supply in series to each parallel switch was proposed there. Finally, such a multi-purpose configuration permits the use of all the operational modes mentioned so far [17] and it is shown in Fig. 16. Another multi-purpose network Ref. [18] is illustrated in Fig. 17. Using thyristors it is self-commutating and needs only one pulse-power supply. However, the flat regions are uncontrolled.

4 Power Sources

Generally, compensation of the power losses is carried out separately for the DC and AC components and at different parts of the network as mentioned above.

4.1 DC Sources

DC is fed into a place in the network where there is minimum AC and since the DC power supply is mostly closed by an LC low pass filter circuit, its capacitor supports the separation of AC from the DC source. Thus the equipment layout is the same as for any DC magnet circuit, e.g., 6-pulse or 12-pulse thyristor converter. However, the load characteristic slightly differs from that of the usual DC magnet circuits. Figure 18 shows the phase and magnitude of the frequency response of the magnet current versus input voltage for a simulated strongly coupled (k = 1) three-mesh White Circuit described above. The load behaviour is that of a first-order system together with a coupled resonance at the network main resonance frequency with a positive 180° deviation over a small frequency range. The practical effects of this behaviour are not known.

4.2 AC Sources

There is a great variety of power sources for supplying make-up power to the oscillating network. Let us distinguish continuous and pulse-wise AC-generation, equipments with and without DC–intermediate and energy storage circuits, and series and parallel resonant supplies, i.e. voltage and current sources.

4.2.1 Plants with DC interlink

*Pulse Power Supply*

This was the first AC power supply proposed [19]; it consists of a variable DC converter, a low-pass filter $L_L, C_f$ and a pulse choke, forming the pulse branch together with a storage capacitor $C_f$ and a fired thyristor
Th. Its steady-state operation is described in Ref. [20]. It is based on the generation of a half-sinusoid current pulse which is fed into the White Circuit via the parallel auxiliary windings of the main choke(s), thus acting as a parallel resonance circuit. The pulse frequency \( f_p \) is considerably higher than the main resonance frequency \( f_0 \). The current pulse is supplied when the main circuit voltage is at its maximum, and the pulse duration is symmetrical to this maximum (or to the magnet current crossover). In this case, the peak voltage of the main circuit \( U_e \) will always adjust itself very closely to the DC supply voltage, see Fig. 19, therefore \( U_{cm} \approx U_0 \).

The size of the storage capacitor \( C_f \) is optimum if it is just discharged to zero voltage by the current pulse. Then its peak voltage is \( 2U_0 \). The behaviour of the pulse current \( i_f \), the storage capacitor voltage \( U_{cf} \), the filter choke voltage \( U_{cf} \) and the filter current \( i_f \) are described by the equations of Annex 1 for steady-state operation. The line diagrams of Fig. 19 represent its mode of operation for \( \frac{\omega_f}{\omega_0} = 3 \) and \( \frac{\omega_f}{\omega_0} = \frac{1}{4} \). With these values the following current ratios are obtained:

\[
\frac{i_p}{I_{pev}} = 8.98; \quad \frac{i_f}{I_{fev}} = 1.11
\]  \hspace{1cm} (22)

In Ref. [21] the dynamic behaviour is analysed. For frequencies much smaller than the basic accelerator frequency \( \omega_0 \) the transfer function \( \frac{U}{U_0} \) (DC magnet voltage : AC supply voltage) is one of second order. Table 2 indicates the harmonic content of the pulse power supply output for \( \frac{\omega_f}{\omega_0} = 3 \) and that of the the magnet current for a resonant circuit quality \( Q_r = 100 \).

<table>
<thead>
<tr>
<th>Table 2: Pulse Power Supply – Harmonic Content</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \nu = 2 )</td>
</tr>
<tr>
<td>PPS output ( \frac{I_f}{I_{pe}} ) =</td>
</tr>
<tr>
<td>Magnet current ( \frac{I_{m}}{I_{m_0}} = 3 \cdot 10^{-3} )</td>
</tr>
</tbody>
</table>

**Continuous Operating Converters**

All types of continuous operating converters with DC interlink can be designed for one of the two inverse operation modes:

- as voltage source DC converter with a storage capacitor as output element (which forms an LC-low pass filter together with a filter choke), feeding into a series resonant circuit

- as current source DC converter with a storage choke as output element, feeding into a parallel resonant circuit.

Between the DC converter output and the load an inverter acts as switching device. A parallel resonant
circuit is formed by connection to the paralleled auxiliary windings of the the main choke(s) as for the pulse power supply. A series–resonant circuit requires that additional transformers be placed in series to the combined capacity $C_{11} + C_{21}$ in each group of the White Circuit as shown in the basic circuit diagram of Fig. 1. All following converters with DC interlink and continuous operation have the same general structure, namely:

- back-to-back arrangement, controllable valves/diode for voltage sources
- only controllable valves for current sources

Resonant Circuit Inverter [5,22,23]

The inverter is a load-commutated type if thyristors are used. This means that after crossing zero the voltage of the thyristor being extinguished has to be reversed for a sufficient duration to ensure that it blocks sufficiently. That is done by a capacitive phase shift of the load, which is therefore operated as a series–resonance circuit below its resonance frequency and as a parallel–resonance circuit above it.

Energy supply for commutation is independent of this. It is delivered from the load in the case of the parallel–resonant inverter (a) but from the DC source in the case of the series–resonant inverter (b). Figure 20 shows the equivalent circuits of both type a and type b and current and voltage wave shapes of type b. In the ideal case, i.e. $L_f, C_f \rightarrow \infty$ and the quality of the load $Q$ also very high, the inverter delivers a square–wave voltage resp. current, and the current resp. voltage of the load is sinusoidal. But due to the limited $Q$–value the current resp. voltage is composed of portions of damped sine curves and due to the finite sizes of $L_f$ and $C_f$ the load reacts to the DC intermediate circuit [24] and increases the share of harmonics of the inverter output.

Both effects cause odd harmonics in the magnet current which may be too large in some cases and make it necessary to insert filter circuits between power supply and load. Since the whole system is to be understood as a self–excited oscillator, the presence of conventional low–pass or resonance filter networks may influence the natural frequency. This can be avoided by careful damping. Another method of decreasing harmonic components is the series connection of the parallel–resonant magnet circuit with a series–resonant LC–filter circuit having nearly the same resonance frequency [25,26]. As this combined circuit must have a series–resonance characteristic for satisfactory operation of the inverter, the filter elements have to be of considerable size in relation to the size of the main circuit, let's say 5 – 10 % expressed in kVA at the accelerator frequency. Due to this supplement the input current fed into the main circuit approaches sinusoidal instead of rectangular shape but a more complicated network, especially for the servo–loop system also has to be considered.

In most cases the conditions for commutation are not fulfilled when the system is started with zero energy stored. Therefore resonant–circuit inverters need a starter device, the simplest possibility would be a temporary connection between the DC rectifier output and the main circuit.
Regarding dynamic behaviour, the frequency response of the load of the resonant–circuit converter without any filter devices has a second-order characteristic of the ratio input inverter valves to magnet current. Amplitude control is possible with this system, but it requires a change in output frequency. Assuming high Q-values, frequency change is small if only a small percentage of the amplitude control is achieved with the inverter and the larger portion by the controllable DC rectifier. Of course, current regulation in this manner has to take into consideration the minimum phase angle required of the converter.

Accelerators with separate function magnet design require tracking of several magnet circuits. In this case the accelerator frequency has to be the same for all circuits, and for current control only the DC converter is available. For this, the parallel–resonant converter is superior because of its better dynamic characteristic: the transfer function DC supply voltage: magnet current represents a system of third order instead of fourth order as in the case of the series–resonant system.

**PWM Inverter**

This inverter is based on the same principal scheme as for the resonant–circuit inverter, see Fig. 19. In contrast to this it has the ability to interrupt the current at any instant, e.g. by use of forced commutated thyristors or GTOs, and is therefore not dependent on the production of output in rectangular shape only. With a suitable pulse pattern it is possible to suppress the harmonic components of lower order [27]. Thus with only one block per half period extended between $30^\circ$ and $150^\circ$ the third harmonic is eliminated. N pulses per half-period makes it possible to eliminate all harmonics $\nu$ in the range $3 \leq \nu \leq 2N - 1$ and if placed in an appropriate manner this is valid over the whole control range of the fundamental component [28]. In Fig. 21 a pulse pattern for $N = 5$ and for 100 % and 60 % fundamental is shown. However, with this method the harmonics problem is only shifted to greater ordinal numbers, the higher harmonic components are increased and could exceed allowed tolerances. In these cases a compromise should be preferred, so that small amounts of lower harmonics are tolerated in favour of the reduction of higher ones. In addition to this, the diminution of the pulse duration could help. Furthermore, there are other pulse patterns to suppress harmonic components and finally a method with fixed pulse periods is applied, which is also described in Fig. 21 b.

The PWM inverters are for four–quadrant output operation. During the gaps between the pulses, the load is bypassed by the flywheel diodes in the case of a voltage source, whereas the inverter input is shortened by two controllable valves in the case of a current source. The latter is also capable of inverter operation DC–interlink — mains with a DC converter only for one current direction. This, together with the simpler connection to the load circuit, is the reason why this inverter type is superior. Two points still have to be considered for this solution:

- to keep a second order load characteristic for dynamic operation the series inductance between inverter and load, e.g. stray reactance of a transformer or main inductor, has to be small;
• $\frac{d}{dt}$ has to be restricted with regard to the allowed values of the thyristors, GTOs, etc.

As the size of the filter elements of the DC intermediate circuit is not infinite, there is a load current reaction similar to the resonance circuit inverter. Low output frequencies increase this effect and a certain amount of active power swing remains. The given tolerances determine the size of the filter elements.

4.2.2 Plants without DC Interlink

These converters are voltage sources and convert AC directly to AC. Thyristors used in such equipments are operating under line-commutation. From there the structure and function is that of a multipulse — 6 or 12 pulse. — DC converter but, in contrast, control is achieved in a time-variable manner. The output frequency must be significantly smaller than the line frequency of 50 or 60 Hz.

*Modulated Rectifier*

In this case, a combined DC–AC source feeds into a series connection of magnets $L_1$ with the whole circuit capacity $C_1 + C_2$, the latter paralleled by the choke inductance $L_2$ [29]. This forms a two-quadrant converter, supplying unidirectional current and positive and negative voltage. Interchanging the positions of the magnets and choke improves the suppression of the magnet current's harmonic and, if $\alpha = \frac{L_1}{L_2} < 1$, the source always produces positive current, even if the magnet current becomes slightly negative before the time of injection [30].

A scheme of the principal is shown in Fig. 22 with typical wave forms in Fig. 23. The combined DC–AC power supply has to be designed for a higher rated power $P_x$ than two separated converters; for DC having the power $P_{DC}$, and for AC (here as rectifier with storage elements supplying DC in an intermediate circuit, followed by an inverter), having the power $P_{AC}$.

Introducing $\dot{U}_{AC} = \kappa \cdot U_{DC}$ and $\alpha = \epsilon = 1$, we get:

$$P_{DC} = U_{DC} \cdot I_{DC}$$ (23)

$$P_{AC} = \frac{\kappa}{2} U_{DC} \cdot I_{DC}$$ (24)

$$P_{C} = U_{DC}(1 + \kappa) \cdot 2 \cdot I_{DC} \cdot \frac{\sqrt{5}}{8}$$ (25)

$$\frac{P_x}{P_{DC} + P_{AC}} = 1.224 \frac{1 + \kappa}{1 + \frac{\kappa}{2}}$$ (26)

A combined converter covering 175 % power ratio compared with two single rectifiers combined to give 100 % power does not cost 75 % more money, of course! More important are the following properties:

• the frequency response of the supply–voltage to magnet current is of third order. This requires a more complex servo–loop system, but gives good harmonics suppression.
• Very simple handling because this power supply is similar to usual equipments feeding DC magnets.

• Fluctuating active and reactive power. The latter can be compensated but is hardly possible because of the influence of the active power swing on the phase angle of the line voltage [31]. Therefore a modulated rectifier is restricted in size if there is no second independent power net from which it is supplied and the consumers of which are insensitive to such variations.

**Cycloconverter**

The difficulties caused by the fluctuating power are removed if the AC power source is made with a three-phase output of the generated frequency. Although only one phase is needed for the magnet resonance circuit, three phases are generated and the "useless" ones are equipped with the reactive power elements of the well-known Steinmetz Circuit [32]. With this method, a one-phase active load having the current \( I \) is compensated for if a capacitive current \( \frac{I}{\sqrt{3}} \) is flowing in the second phase and an inductive current \( \frac{I}{\sqrt{3}} \) in the third one. Active and reactive loads are delta-connected, and the kVA-size of the three-phase converter is \( 3 \cdot \frac{I^2}{\sqrt{3}} \cdot \frac{I}{\sqrt{3}} = UI \) and has the same value as a one-phase converter feeding directly into the resonance network. If the magnet load represents a parallel resonant circuit, inductors have to be put between the cycloconverter and the three-phase load, see Fig. 24. In this case the AC voltage source has to deal with a third-order system as regards the dynamic behavior from supplied voltage to magnet current. Two types of cycloconverter operation are known: with and without circulating current. Figure 24 deals with the latter and Fig. 25 shows output current and voltage curves. Its structure is simpler because reactors which limit the circulation are not necessary, in contrast to the other type. However, a zero-current interval is necessary to make sure there is enough recovery time for the thyristors. Of course, such a converter type cannot be used for a series-resonant arrangement of the magnet network.

Adequate control methods should eliminate the influence of the line frequency on the intermission duration: due to the fine structure of the generating voltage, harmonic currents which depend on line frequency can influence the load current zero crossover and thus cause a frequency modulation if the magnet cycle is not strongly coupled to the mains frequency. The use of a cycloconverter with circulating current might be an alternative.

The cycloconverter, too, has to be designed for a larger power size compared to a usual DC-converter on the basis of equal output power. Thus the ratio of the transformer sizes is 1.41 [33].

### 4.2.3 Protection and Tuning

In addition to the normal protection methods used for converters, care must be taken that in the case of any interruption:

• the continuation of current flow in an inductor is ensured

• capacitors are not overloaded.
This is obvious in the case of a series-fed resonant circuit. Here interruption may only occur when the whole multi-mesh network is de-energized if overvoltages by addition of capacitor voltages are to be prevented. Protection for voltage source supplies with DC interlink or unidirectional output only requires one thyristor fired by the alarm signal. In practice, however, one can avoid the short-circuit of a voltage source by a protection thyristor and rather bypass bidirectionally series-resonant circuits in case of faults directly, e.g. by one type of electromechanical device, which has the reputation of operating fast and very safely. In this case, the load circuit has to be separated by an inverter in open state. Pure parallel resonant circuits are self-discharging. Therefore, protection is only necessary if there is series inductance. The alarm and ignition circuits must be independent of the mains.

In most cases magnet operation with fixed frequency is desired. Then, tuning is needed because of the temperature drift of the main capacitors and the relatively high Q-value of the resonant network, which varies between 10 and 300 depending on the accelerator size and frequency.

Between extreme ambient temperatures, the capacitance variation of capacitor banks of normal all-film type units installed outdoors, is 3%. Tuning is done by switching capacitors in each group in rough steps and fine tuning by a choke controlled by back-to-back thyristors, as shown in Fig. 24. This allows fine tuning while avoiding too frequent switching. The variable choke has proved to be a very useful instrument for quick and precise tuning, e.g. in the following cases:

- to adjust the dynamic characteristics of different resonant circuits for tracking operation (e.g. dipole and quadrupole circuits)
- to adjust the magnet resonant circuit to the line frequency in a defined manner to avoid beats of difference frequency

The Steinmetz compensation is sensitive to mistuning. Fortunately, the magnet circuit load is represented by a constant resistance and therefore one also has to take care that the compensating elements are constant. If tuning is required in spite of this, the control system for this operates independently of the control system of the magnet current.

4.3 Automatic Control

Generally the control methods depend on the type of the AC power supply and a distinction can be made between power sources with only one regulating unit, i.e. all power supplies without DC interlink, and power sources with two regulating units, i.e. all equipments with DC interlink.

Figure 26 shows a control system of the first mentioned type for the control of a combined AC-DC supply (modulated rectifier). The three variables controlled for AC amplitude, AC phase angle and DC are all "DC"-quantities. Since the AC amplitude is defined and measured twice per period, stabilization of the AC amplitude does not produce a sufficient result if the Q-value of the resonant circuit is low as for
low magnet frequencies. Help from a subloop of the actual unfiltered voltage of the power supply output improves the tolerances because a step in line voltage is the only fast disturbance variable.

The second type deals with the control of a pulse power system, see Fig. 27. Here, two feedback systems are necessary: one for the pulse frequency resp. phase angle which affects the pulse thyristor, and the other for the control of the magnet current amplitude.

Timing and amplitude control may be combined as in the case of the PWM-converter. Here a single fast feedback loop working on the inverter part is able to control the magnet current amplitude as well as its frequency and the dynamic system belonging to it is only of second order. The DC converter should only be used for follow-up control to limit the correcting range of the inverter.

So far, the actual value of the magnet current changes stepwise with step distances of at least one period. There are two possibilities of avoiding the difficulties inherent in this method and for producing a continuous change:

- a process consisting of three stages: demodulation of the AC (suppression of the carrier), application of feedback control methods and modulation with the carrier [34]. This method is well suitable for a three-phase system and therefore for the input currents of the cycloconverter with three-phase output, as mentioned above.

- Measurement of the instantaneous AC magnet current and comparison with the AC reference value (oscillator). The controlled variable is their difference, is an AC quantity and is processed in the servo-loop system [35]. Thus the oscillator moved from the inverter – or converter – input to the controlled variable. The current feedback loop has no integrating network as for DC loops because integration would only result in a 90° phase shift. This part is replaced by a resonant network for the magnet fundamental frequency based on the following idea [35]:

\[
\sin \omega t
\]

deviation of the controlled variable:

\[
k t \cdot \sin \omega t
\]

output signal of the current feedback loop after time t:

\[
\frac{\omega}{p^2 + \omega^2}
\]

the Laplace transforms are:

\[
\frac{2 \ p \ \omega \ k}{(p^2 + \omega^2)^2}
\]
The quotient $(30):(29)$ represents the transfer function required of the resonant network.

\[ \frac{2p \cdot k}{p^2 + \omega^2} \]  \hspace{1cm} (31)

In the real case the deviation of the controlled variable is not purely sinusoidal but has side-band frequencies. Therefore $(31)$ is replaced by a transfer function of a weakly damped band pass filter. Figure 28 shows the effect of such a device on the input function $\left(1 - e^{t/T}\right) \sin \omega_0 t$. The transfer function is produced by reversed operational amplifiers having RC-networks [35].

5 Conclusion

It has been demonstrated that resonant excitation offers a large variety of methods and possibilities. The following comments might be helpful in making a choice:

Joint inductors reduce, of course, the size of the installed material and, as shown above, the electric losses but need a very special design. For example, the number of turns of the main and equalizer coils of the window frame type has to be calculated carefully. In case of a failure, repair work takes longer than for independent chokes where only a complete spare unit has to replace the damaged one. In addition, single chokes can be installed closer to the magnet and capacitor group belonging to it. That means shorter connections and thus less stray capacitance.

Closely coupled and paralleled equalizer windings permit the use of an extended network ring like a lumped LC-circuit and are therefore absolutely necessary.

Superposition of harmonics, especially with dual chokes, offers a soft method in wave shape variations. The necessity of an additional power source for these processes is an advantage - and not a disadvantage as sometimes stated - in producing a further possibility to influence the stabilization of the magnet current. Dual chokes are also a suitable input device for active filters. In contrast to this, switch mode operation generates transients in multi-mesh White networks, if switching does not occur synchronously; transients might also be caused by commutating processes. Dual frequency mode excitation by switching off capacitors causes significantly higher AC voltages - which the magnets have to withstand - than the method with harmonic superposition. Therefore, switch mode operation should only be applied if there is no other possibility, e.g. for a long-time flat-top.

Continuously operating AC power sources should be preferred if at all possible. For repetition frequencies $\leq 10$ Hz, and large power sizes, the cycloconverter with Steinmetz compensation is a suitable device and its behaviour with circulating current feeding a resonant network should be investigated. In the whole field, the current source GTO equipped PWM converter is of great interest.

Parallel-connected power sources are superior to series-connected ones since the fault protection of the whole system is much simpler.

Future improvements should be directed to:

- Power sources and stabilization of the magnet current for controlled switch-mode waveforms, e.g. for long-time flat-tops;
- Continuous working AC magnet current loops in automatic regulation systems.
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Annex 1

Pulse Power Supply Equations

The equations for Periodic Operation [20] of the circuits shown in Fig. 19 using the approximations

\[
\omega_f = \frac{1}{\sqrt{L_f C_1}} \quad \omega_p = 2\pi f_p = \frac{1}{\sqrt{L_p C_1}} \quad C_m \gg C_p; \text{ no losses} \quad U_{ef} = 2U_o
\]

and a basic accelerator frequency

\[
\omega_0 = \frac{1}{2\pi f_0} \approx \frac{1}{\sqrt{(C_1 + C_2) \frac{L_0 L_1}{L_0 + L_1}}}
\]

are:

1) Pulse period

\[
-\frac{1}{4f_p} \leq t \leq \frac{1}{4f_p}
\]

Pulse current

\[
i_p = \frac{U_0}{\omega_f L_f} \left[ \omega_p \cos \omega_p t + \cot \frac{\pi}{2} \left( 2 \frac{\omega_f}{\omega_0} - \frac{\omega_f}{\omega_p} \right) \left( 1 + \sin \omega_p t \right) \right]
\]  \hspace{1cm} (A_1)

Pulse current - peak value

\[
i_p \approx \frac{U_0}{\omega_f L_f} \left[ \cot \frac{\pi}{2} \left( 2 \frac{\omega_f}{\omega_0} - \frac{\omega_f}{\omega_p} \right) + \frac{\omega_f}{\omega_p} \right]
\]  \hspace{1cm} (A_2)

Filter current

\[
i_f \approx \frac{U_0}{\omega_f L_f} \left[ \cot \frac{\pi}{2} \left( 2 \frac{\omega_f}{\omega_0} - \frac{\omega_f}{\omega_p} \right) - \frac{\omega_f}{\omega_p} \cdot \cos \omega_p t \right]
\]  \hspace{1cm} (A_3)

Storage capacitor voltage

\[u_{C_f} \approx U_0 \left( 1 - \sin \omega_p t \right)\]  \hspace{1cm} (A_4)

Filter choke voltage

\[u_{L_f} \approx U_0 \cdot \sin \omega_p t\]  \hspace{1cm} (A_5)

2) Charging period

\[
-\frac{1}{4f_p} \leq t \leq \frac{1}{f_0} - \frac{1}{4f_p}
\]

Filter current

\[
i_f = \frac{U_0}{\omega_f L_f} \frac{\cos \left( \omega_f t - \pi \frac{\omega_f}{\omega_0} \right)}{\sin \frac{\pi}{2} \left( 2 \frac{\omega_f}{\omega_0} - \frac{\omega_f}{\omega_p} \right)}
\]  \hspace{1cm} (A_6)

Storage capacitor voltage

\[u_{C_f} = U_0 \left[ 1 + \sin \left( \frac{\pi}{2} \cdot t - \frac{\omega_f}{\omega_0} \right) \right] \]  \hspace{1cm} (A_7)

Filter choke voltage

\[u_{L_f} = -U_0 \frac{\sin \left( \omega_f t - \frac{\omega_f}{\omega_0} \right)}{\sin \frac{\pi}{2} \left( 2 \frac{\omega_f}{\omega_0} - \frac{\omega_f}{\omega_p} \right)}\]  \hspace{1cm} (A_8)
Fig. 1 DC-biased resonant excitation; basic circuit diagram

\[ L_1 \text{ inductance of the series connected magnets} \]
\[ L_2 \text{ DC bias choke} \]
\[ C_1, C_2 \text{ capacitance, resonating with } L_1 \text{ resp. } L_2 \]
\[ L_2 = \frac{1}{\sqrt{C_1}} = \frac{1}{\sqrt{C_2}} \]

Fig. 2 White network; principle and section

Fig. 3 DC-biased resonant excitation
\[ \alpha = 1, \beta = 0.5 \]
\( a \): voltage, currents
\( b \): stored energy

Fig. 4 Bypass parallel resonant circuit
Related costs \( C_r = f(\alpha) \), Parameters \( \beta = \frac{C_r}{C_C}, Q_{\text{in}} \)
Fig. 5 CEA-twelvefold ring choke [3]. Section of 3 parts

Fig. 6 DESY-twelvefold compound inductor
stored max. energy $W_a = 1.46 \cdot 10^4$ Ws
Fig. 7 3-mesh White Circuit – basic circuit diagram and voltage distributions

a fundamental, undisturbed  
b due, with earthfault  
c, d additional voltage components  
of spurious mode resonances due to asymmetrical coupling [8]

Fig. 8 3-mesh White Circuit  Spurious modes
Fig. 9 White Circuit  Delay line modes and their suppression [8]

a: equivalent network  b: magnet current = fundamental mode

c: damping loop

Fig. 10 Superposition of harmonics

Fig. 11 Superposition of harmonics by resonant circuits

Example
Fig. 12 Superposition of harmonics by dual chokes — mode of operation

Fig. 13: Switch–Mode Flat–Top with additional Chokes

- Basic circuit diagrams
  - a: circuit diagram for short time flat–top
  - b: circuit diagram for long time flat–top

Fig. 14: Long time flat–top with additional chokes

waveforms for \( a = 1; \ v = 1 \)
Fig. 13: Dual frequency magnet excitation by capacitor switching

Fig. 16 Multi-purpose switch mode network [17]
Fig. 17 Network for uncontrolled flat-top and flat-bottom modes [18]
Fig. 19  Pulse power supply [20]

Equivalent Circuit and waveforms for $\frac{V_p}{V_o} = 3$, $\frac{V_F}{V_o} = \frac{1}{4}$

Fig. 20  Parallel -a- and series -b- resonant converters

Equivalent circuits and waveforms for b.
Fig. 21 PWM-inverter outputs

Fig. 22 Modulated rectifier - equivalent circuit
Fig. 23 Modulated rectifier - output and input quantities

Fig. 24 Cycloconverter with 3-phase output and Steinmetz Compensation

Fig. 25 Cycloconverter - output quantities

Input: 12-pulse, 50 Hz  Output: 12.5 Hz
Fig. 26 Modulated rectifier – multi-loop control circuit

\[ \Delta I_m \rightarrow V \rightarrow \alpha \rightarrow \text{Oscillator} \rightarrow X \rightarrow \text{Gate firing set} \rightarrow \text{Gate firing pulses} \rightarrow \alpha \phi \rightarrow I_{ac} \rightarrow I_m \rightarrow I_{ref} \]

\[ \Delta I_m = I_m - I_{ref} \]

Fig. 27 Pulse power supply – multi-loop control circuit

\[ V = \frac{1 + p T}{1 + 2 \lambda_0 \beta (\lambda_0)^2} \]

\[ T = 2 \pi, \beta = 0.02, \omega_0 = 2 \pi \times 12.55, d = 0.12 \]

Fig. 28 Frequency and time response of a band pass filter circuit used as servo-loop component for continuous recording of the AC output quantity.

\[ T_c = \frac{30}{\omega_0} \]
FAST-PULSED CONVERTERS

J. Rümmler

ABSTRACT
In order to become familiar with the use of septa and kickers with their associated pulsers within the DESY network (Fig. 1), the DESY injection chain is discussed by way of introduction. Kicker pulsers often require the attention of specialists. A general overview of the types of pulser is presented, leading to a more detailed treatment of a reversible-polarity PFN pulser for short pilot pulses and rectangular pulses. Pulser design must be based on the kicker magnets to be excited and the effect which they have on the accelerator beam.

Fig. 1 DESY accelerator network

1. SEVERAL DIFFERENT PULSER TYPES

1.1 Kickers and half-wave pulsers in the PIA ring

Fast kickers are used here to produce a beam bump. They direct the injected beam to the vicinity of the septum blade, in order that the circulating and injected beams are as close to each other as possible (within the tolerance of the machine) after one revolution. Depending on the machine, the beam bump generated by the kicker must be switched off quickly, otherwise it is possible that the beam will be scraped off at the septum. The location of the kickers in the Positron Intensity Accumulator (PIA), together with their constructional details are shown in Figs. 2 and 3.
Fig. 2 Location of the kickers in PIA

Fig. 3 PIA kicker and pulser
Kickers and pulsers of the same type are used for ejection. In PIA the particles are collected to form a bunch before ejection and, at the instant of ejection, are switched into the ejection channel via the septum magnet. The kicker pulse therefore has one bunch revolution for its pulse risetime. For this reason, a half-wave pulser with a low internal resistance is employed here. Rectangular pulse kickers excited from high impedance pulsers cannot be justified because of their increased cost.

1.2 Injection kickers and pulsers in DESY 2

Fast half-wave pulsers are used with toroidal tape cores as a quenching aid for the thyratrons. In the circuit shown in Fig. 4, toroidal tape cores are shown at LB. The Z characteristics of the cores make them ideal switches between the states of high saturation and very high inductance (even at currents as low as 2 A). When the current crosses the zero axis in the thyratron, a voltage divider is created. High voltage appears across the toroidal tape core lying between point A and the anode, and C2 keeps the voltage across the thyratron low, so that it can quench. Triggering takes place with the choke saturated and without pulse delay. The magnetic flux of the choke in volt-seconds indicates the voltage-time integral held by the choke.

![Fast half-wave pulser circuit](image)

Fig. 4 Fast half-wave pulser circuit

1.3 Ejection kickers and pulsers in DESY 2

If the remanent fields of many kickers interfere with the beam at a low injection energy, ejection kickers under damped pulses can be employed. Simply by using thyratrons with hollow anodes in the pulser, or connecting fast diodes in parallel, the pulser is capable of oscillating as shown in Fig. 5.

![Oscillating kicker pulse](image)

Fig. 5 Oscillating kicker pulse
1.4 Ejection kickers and pulsers in DESY 3

If entire bunch trains are to be ejected from machines, the kickers require rectangular pulses in order to excite the same deflection field for all bunches. Rectangular pulses by cable discharge are the easiest to implement. As an important quality for kickers and pulsers, the characteristic impedance appears for the first time. From transmission-line theory it is known that only line sections with the same characteristic impedance are able to transfer pulses with their entire frequency spectrum and without reflections, and therefore without great pulse distortion. A typical arrangement is shown in Fig. 6.

Thyratron
CX 1168

\[ \text{Schematic diagram of pulser} \]

1.5 Excitation kickers and pulsers in DESY 3

These kicker pulsers operate with two thyratrons, as shown in Fig. 7. At the remote end of the storage cable there is a diode dump. Injection pulsers are also used in this circuit, since thyratron 2 makes the trailing edge of the pulse steeper. The circuit is shown in Fig. 8 where it can be seen that if thyratron 1 is triggered, a complete rectangular pulse with double cable travelling time is created in the terminating resistance. If thyratron 2 is switched on, the pulse is shortened and thyratron 2 reflects the rest of the pulse with phase reversal into the diode dump. Thyratron 2 conducts twice the pulse current, which reduces its life expectancy.

\[ \text{Kicker pulser in the centre of the accelerator ring} \]
1.6 **Pulse-forming networks (PFN's) as pulse drivers**

Kicker pulsers for pulse lengths greater than \( t = 5 \mu s \) are best operated with lumped element PFN's, since long cables tend to distort rectangular pulses and cable losses sometimes generate unacceptable pulse sag. Depending on the pulse slopes and lengths required, the PFN's are designed with various divisions, or numbers of elementary cells, and various characteristic impedances.

As is well known from transmission-line theory, a line with a real characteristic impedance and terminating resistance is able to transfer the frequency spectrum of a rectangular pulse. The PFN design therefore begins with the determination of the resistance for the PFN, cable, kicker and terminator. In the examples, unwanted inductances, losses and capacitances are neglected. Unwanted inductances can, however, distort rectangular pulses quite considerably.

The pulse length of the PFN is given by

\[
\tau = 2\sqrt{L \cdot C}
\]

and the characteristic impedance is

\[
Z = \sqrt{L/C}, \quad \text{typically 10 - 20}\,\Omega.
\]
Fig. 9 PFN without front cell
Fig. 10 PFN with front cell
Through its resonance spectrum, the front cell, consisting of \( R_f, C_f \), and \( L_f \), improves the curve shape of the rectangular pulse to a large degree. As an alternative, the PFN can be built at lower cost with fewer elementary cells. In Figs. 9 and 10 curves are shown for PFN's with and without a front cell. With PFN's of higher resistance, small reflections have a minimal effect. Thus, PFN's used with kickers are not usually designed for less than 10 \( \Omega \).

1.7 PFN pulsers for injection kickers

PFN's or cable discharge can be used in a similar manner as pulse drivers, and so PFN's are also suitable for pulsers of variable pulse length (see section 1.5).

1.8 PFN kicker pulsers for long pilot pulses

The proven PETRA ejection to HERA is used as an example. As often stated at DESY, the boundary conditions for this pulser type are based on PETRA operation. For this, ejection requires three kicker magnets with three pulsers whose main requirements are:

- rectangular pulses for ejecting all bunches (\( t < 7.6 \mu s \)) in a bunch chain to HERA.
- ejection at various energies with variable high voltage at the pulser.
- PETRA operation, with 7 GeV injection and subsequent ramp operation, high-power acceleration up to storage at the ejection energy, requires short pulses or pilot pulses to eject single bunches from the stored beam into the transport path and to HERA. This must not interfere with the beam remaining in PETRA for the actual ejection.
- Reversibility of the kicker pulser because either electrons or positrons are ejected from PETRA. In order to reverse quickly, dual-cathode thyatron systems are used. Depending on the polarity of the power supply, and thus the charge of the PFN, electrons or positrons are ejected.

The main features of the system design with the above boundary conditions are:

- Due to the long pulse (\( t = 7.8 \mu s \)), the pulser is driven by a PFN.
- In order to avoid using a pulser under oil, a low characteristic impedance of 10 \( \Omega \) is selected for the required pulse slopes.
- The kickers are divided into RC elements with the somewhat larger capacities outside the kicker tank.
- Dual-cathode thyatron systems permit fast reversal for positive or negative pulses without generating unwanted inductances in the pulser.
- With three thyatron systems, the pulser makes it possible to have pilot pulses and fast reversal.
- A chopper-regulation power supply with triggered stop and go is used. The fast regulation is state of the art and regulates much faster than is possible at 50 Hz. Recharging takes place without an impedance or separating choke to the pulser. Power supply and pulser are located about 100 m from the kicker.
- The PFN pulser has an impedance of 10 \( \Omega \).

Some of the above features are illustrated in Figs. 11-15.
Fig. 11 Three kickers and pulsers for the PETRA ejection

Fig. 12 Long and short pilot pulses for the kickers
Fig. 13 The kicker of the PETRA ejection is driven by the PFN
Fig. 14 The kicker in the PETRA ring tunnel

Fig. 15 PFN and pulser
PULSED CAPACITOR DISCHARGE POWER CONVERTERS: AN INTRODUCTORY OVERVIEW.*

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ABSTRACT
The applications where pulsed power converters are used are recalled and the converters classified according to the specificity of their basic constituent power parts and electronics. Present technical solutions are described and development trends mentioned while specialized reports are referred to for more technical details.

1. INTRODUCTION

Under the term ‘pulsed power converters’ one understands the family of apparatus based on the charge-discharge of either lumped element Pulse Forming Networks (PFN) or of capacitor banks, designed to produce current pulses with a duration of up to several ten ms and an amplitude of over 1.2 MA.

These power converters are used, instead of dc or programmed converters, to achieve high magnetic fields in very compact deflecting or focusing devices and to perform particular time dependent current patterns for accelerator operations, as shown by the following non-exhaustive list of possible applications (usual current levels are given to characterize the equipment):

- Beam transport bending and quadrupole magnets (I < 3 kA)
- Correction and steering dipoles (< 100 A), [1]
- Transversal beam position scanning dipoles (< 200 A)
- Switching magnets to select final beam destination (< 3 kA)
- Switching magnets for beam distribution or recombination (< 1 kA)
- Magnets for local excitation of beam oscillations for ejection (< 2 kA)
- Quadrupoles for gamma transition schemes (< 2 kA)
- Thin monotum septum magnets (< 50 kA)
- Current carrying targets for p- production (< 0.5 MA)
- Coaxial lenses to collect neutrino parents or p- behind a target (< 0.5 MA)
- Lithium and plasma lenses either to focus the primary p beam in front of the target, or to collect p- or for final focus schemes (< 1.2 MA)
- Super strong quadrupoles for colliding beam focusing (< 0.5 MA) [2]
- Particle collecting and matching solenoids behind e+ production targets (< 20 kA)
- Special radiation-hard post target magnets (< 100 kA)
- Complete fast cycling ultra-compact accelerators and beam extraction gantry systems for medical and industrial applications (< 25 kA) [3,4].

The advantages of pulsed power converters (high efficiency and reduced power consumption, compact magnets, possibility of pulse-to-pulse current modulation in time and amplitude, possibility of multiple pulsing, fast rise time and precise current flat-top regulation) are to be compared with some penalizing features (pulsed low

* This topic was not presented at the School but is included here to widen the scope of the syllabus.
cos \( \phi \) power demand from the mains, higher ac current harmonics content, need of accurate timing pulses to control the charge-discharge sequence, less conventional technology and mode of operation). Nevertheless, thanks to the high adaptability of designs for an ever increasing number of applications, pulsed power converter technology has accompanied the evolution of particle accelerators and will certainly play an even more important role in the frame of future very high energy linear lepton colliders.

2. COMPOSITION OF PULSED CONVERTERS FOR CAPACITOR CHARGE, ENERGY STORAGE AND DISCHARGE

To classify the pulsed power converters one can consider them as consisting of a number of basic functional subassemblies, namely:

i) a mains fed energy supply and charging circuit;

ii) an energy storage PFN or capacitor bank, possibly with third harmonic current pulse shaping;

iii) a discharge circuit, possibly including an energy conversion or recovery unit, an active filter for current flat-top regulation, a pulse transmission line and a load impedance matching transformer;

iv) electronics to fulfil the control, monitoring, timing and regulation functions.

2.1 Charging circuit

The type of charging circuit depends both on the time available and on the power level of the pulsed converter. As indicated in Table 1, the most common methods are resonant capacitor charging (Fig. 1) [5], use of a higher frequency chopper (Fig. 2) and linear charging (Fig. 3) [3]. In general for charging times \( > 0.3 \) s the mean value of the charging current is kept constant for linear charging of the energy storage element. More recent designs foresee charging at constant active input power to alleviate mains loading, especially in the case of higher power ratings.

<table>
<thead>
<tr>
<th>Charging time (s)</th>
<th>Implemented charging circuit solutions</th>
</tr>
</thead>
<tbody>
<tr>
<td>(&lt; 0.05)</td>
<td>Resonant charging with controlled charge interruption or 'deQuing'</td>
</tr>
<tr>
<td>(0.05 - 0.3)</td>
<td>Higher frequency chopper</td>
</tr>
<tr>
<td>(&gt; 0.3)</td>
<td>6-pulse or 12-pulse (if (&gt; 200) kW) thyristor controller on primary of stepping-up transformer</td>
</tr>
</tbody>
</table>

2.2 Energy storage circuit

Lumped element PFN's (\(R_o = 25\) Ohm) have been designed to produce current pulses of up to 200\( \mu \)s duration with rise and fall time better than 0.1\( \mu \)s at a repetition frequency of 1 Hz (Fig. 4) [6]. They require special low-inductive pulse capacitors.
Fig. 1 Power converter with resonant charging within 8 ms for e-+ converter solenoid of the LEP injector

Fig. 2 Power converter with higher frequency chopper charging for the storage ring injection septum magnets at ESRF (presently under construction)

Fig. 3 Converter with linear charging through thyristor controller on primary of transformer rectifier assembly for the AAC lithium lens pulser
When approximately sinusoidal current pulses are required, simple capacitor banks are used. The most common energy storage capacitor for pulsed applications is the mixed dielectric type (plastic film, paper) with aluminium armatures and either natural (mineral or castor oil) or synthetic oil impregnation. These capacitors are specified as industrial 50 Hz ac units with appropriate ratings to simplify their procurement [7]. In certain cases the capacitor bank is subdivided into two parts, or an extra parallel LC circuit is added to it, in order to superpose a third harmonic component of given amplitude to the basic sinusoidal discharge current (Fig. 5).

Concerning the maximum stored energy, which is kept to about 20 kJ per cubicle for reasons of industrial safety, a power converter has been recently built with a capacitor bank of 200 kJ for the pulser of the p- collecting lithium lens (see Fig. 3) [8]. A tentative classification of the energy storage circuits is shown in Table 2.

<table>
<thead>
<tr>
<th>Voltage level</th>
<th>Stored energy</th>
<th>Type of capacitors</th>
<th>Third harmonic</th>
</tr>
</thead>
<tbody>
<tr>
<td>U &lt; 1 kV (LV)</td>
<td>E &lt; 1 kJ</td>
<td>Industrial &lt; 20 kJ</td>
<td>Not present</td>
</tr>
<tr>
<td>1 kV &lt; U &lt; 10 kV</td>
<td>1 kJ &lt; E &lt; 20 kJ</td>
<td>Special</td>
<td>Separate LC (Fig. 5 (1))</td>
</tr>
<tr>
<td>U &gt; 10 kV</td>
<td>E &gt; 20 kJ</td>
<td></td>
<td>Integrated LC (Fig. 5 (2))</td>
</tr>
</tbody>
</table>

* N.B. Tables 2 and 3 list different characteristics of pulsed power converters without systematic relation between lines of different columns.

2.3 Discharge circuit

The PFN or the energy storage capacitor bank is discharged into the magnet load by means of thyristors, thyratrons or, more rarely, by igniton switches. Ignitrons have practically been replaced by thyristors while thyratrons are still used where high voltage, high current and di/dt, fast rise time and pulse repetition rate are required.
Fig. 5 Superposition of third harmonic component on main discharge pulse: electrical circuit characteristics and waveforms

A number of different discharge schemes are in use, as shown in Fig. 6, depending on:

- the type of magnet and the degree of magnet current reversal
- the energy recuperation method, i.e. through an auxiliary inductance (1,2), through the magnet load (3) or through the charging choke (4)
- the degree of voltage reversal on the capacitors.

When pulse-to-pulse peak magnet-current modulation is required, the residual energy in the capacitors is dissipated between pulses (if the subsequent peak current value is expected to be smaller than that produced by the voltage after energy recuperation). Double pulsing within the same operating cycle requires either fast recharge of the capacitors by means of an auxiliary circuit or the presence of multiple discharge branches, which are charged in parallel and discharged in sequence.

In applications where a current plateau of $T_p \geq 100 \mu s$ duration and current stability and reproducibility $D/I/I$ better than $10^{-4}$ is required, the energy storage section is equipped with additional third harmonic pulse-shaping components and with a choke for the insertion of an active filter (Fig. 7). This is a MOS-amplifier of class C, with its separate supply and isolated drive, acting in a fast servo-loop and absorbing any peak current variation due to residual ripple of charging voltage or current shape, and to non-compensated thermal drift of the capacitance and of the circuit resistances [9,10].
The discharge circuit can either feed the magnet directly through a pulse transmission line or feed the primary of an impedance matching transformer with a turns-ratio N:1, whose high-current secondary is connected to the magnet. The transformer turns-ratio determines both the discharge current amplitude (I/N) and its pulse duration which is proportional to \(N\pi(LC)^{1/2}\). The transformer has normally a three-limb construction with interleaved windings, located on the central limb, for minimum stray inductance. Either a dc current bias, directly or via an auxiliary tertiary winding, or an air gap of the order of 5 mm are foreseen in the magnetic circuit to cope with the dc component of the excitation current. Transformers with turns-ratio of 6:1, 8:1, 10:1, 12:1, 20:1, 24:1 have been used for pulsed septum magnet applications. The lithium lens pulser (1.2 MA, 4kV) [8] has a 1.5:1 auto-transformer followed by an 18:1 toroidal pulse transformer with dc current bias. A possible classification of the discharge circuits is given in Table 3.
Figure 8 shows the power converter of the most important monotron pulsed septum magnets at the CERN-PS to illustrate the different characteristics of this type of equipment [11]. The energy storage capacitor is divided into two sections for third-harmonic, current-pulse shaping and is charged linearly for pulse repetition times $\geq 1.2$ s. A residual energy dissipation and a fast recharge circuit are added to perform multiple pulsing and pulse-to-pulse current amplitude modulation. A current regulated high stability flat top is obtained by means of a MOS active filter. The septum magnet requires a current of up to 40 kA and a 12:1 matching transformer is inserted between the load and the pulse transmission line.
3. **ELECTRONICS FOR CONTROLS, TIMING, MONITORING AND REGULATION**

3.1 **Controls**

The computer control interface in the CERN-PS is a single transceiver board [10], which can be either fully digital (STD) or have analogue current reference and acquisition (STH-hybrid). Control is based on four 8-bit words for actuation, reference setting, current and status acquisition. The control protocol foresees four exclusive actuation bits: OFF - STANDBY - ON - RESET, and the corresponding acquisition bits. The current reference and acquisition have either 12-bit (STH) or 14-bit (STD) resolution. A number of indicator bits provides information on detailed operational situations (e.g. internal or external interlocks and faults).

3.2 **Timing**

The timing function is an essential aspect in pulsed power converters because they must be synchronized with the accelerator operation in order to achieve maximum stable field in the corresponding magnets when the beam is present. This means that before each discharge one must ensure that the capacitors have been charged to the required level, that the voltage has been stabilized and that all precautions have been taken to avoid short-circuiting of the charging section during the capacitor discharge. Similarly, the discharge must be completed and the switches must have recovered their voltage blocking capability before recharging the capacitors.

To meet these constraints the operating cycle of a pulsed power converter has been subdivided into specific time intervals which are initiated by four timing pulses, at least two of which must come from the external general accelerator timing system. Whenever possible, all four pulses are delivered externally to the power converter for the highest possible operational reliability and transparency.

The four standard timing pulses are:

- **FOREWARNING (FW)** which announces that the power converter is asked to operate during the next beam operation and initiates the charge of the energy storage capacitors
- **WARNING (W)** which blocks the charging circuit once the capacitors have been charged and their voltage stabilized to the required level. This pulse sets the reference to zero and starts all actions foreseen to assure a safe discharge
- **START (ST)** which triggers the discharge so that maximum current is reached in the magnet after a given time interval
- **MEASURE (MEA)** which triggers Sample/Hold acquisition of the discharge current and which normally corresponds to the beam presence and to peak discharge current.

Because of the importance of the timing pulses for successful operation of any pulsed power converter, adequate pulse indication and time interval measuring facilities are part of the normal auxiliary equipment.

3.3 **Monitoring**

The magnet current is measured by means of pulse transformers (e.g., PEARSON, US) or high performance dcct's (e.g., HOLEC, NL or DANPHYSIK, DK). The signal is used for the flat-top regulation loop, and for displaying the current waveform and peak value.
3.4 Regulation

Pulsed power converters with linear charge have, in general, either cascaded or parallel-switched voltage and current control servo-loops. Current control acts during the charging time. After initial soft start and subsequent full current charge, the current reference is reduced for better transients when approximately 90% of voltage is reached. Higher-gain voltage control takes over at transition to the final capacitor voltage level and during stabilisation before the W pulse.

The voltage regulation of resonant charging type power converters is done either by controlling the low voltage source and interrupting the resonant charge as soon as the desired voltage has been achieved (by extinction of the power thyristors) or by keeping the source voltage constant and 'deQuing' the resonant circuit.

A faster and more accurate voltage control is obtained by means of a higher frequency chopper in the charging power circuit. It is possible to profit from this by imposing constant active power demand from the mains during the charging period. For very large power ratings, solutions have been worked out to absorb roughly constant power even outside the charging time (i.e. during the W–FW time interval).

The flat-top regulation by an active filter, when present, constitutes a separate servo-loop with larger bandwidth (100 kHz). In this case the magnet current is directly compared to the reference and the error signal feeds the active filter via a suitably phase-corrected isolated drive circuit. On the base of the power circuit characteristics and by modelling the load, several corrections are applied to the loops to cope with non-linearities between charging voltage and pulse current, with thermal drift due to ohmic losses in the power circuit and with any required time and amplitude pattern of magnet pulse sequences. By these means too large variations of the operating point of the active filter are also avoided during equipment warm-up conditions.

4. CONCLUSIONS

The wide domain of pulsed power converters for particle accelerators has been introduced by means of several examples of recent realisations. The composition of the converters and the technical solutions have been briefly described and references made to more detailed specific technical reports. Trends of development concern the use of advanced power devices (e.g. IGBT's and GTO's) and microprocessors to implement smarter technical solutions and build mains and user friendly pulsed converters.

The combination of a lumped element PFN with an active ripple filter may become an interesting solution for particular applications. Modern semiconductor switches are entering applications where thyratrons were dominating so far and pulse compression techniques with the help of modern magnetic materials are already being applied [12]. More sophisticated operations on present accelerators, and the next generation of high energy linear lepton colliders, will make extensive use of pulsed power converters. Development work in this field should therefore be energetically pursued.
REFERENCES

TOKAMAK PLASMA CONTROL USING THYRISTOR POWER CONVERTERS

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ABSTRACT
Tokamaks are large physics experimental devices for investigating the feasibility of a fusion reactor. High magnetic power is needed to generate and sustain a good plasma. To keep this extremely hot ionized and sometimes very unstable hydrogen gas away from the vessel wall, the magnetic fields have to be controlled very rapidly and accurately with thyristor power converters. For this purpose nonlinear, robust and discrete optimum regulators have been designed and tested. A method is described for finding the best regulator parameters.

1. INTRODUCTION

Nuclear fusion represents one of the options for future energy supply. It is based upon the fusion of the nuclei of deuterium and tritium as illustrated in Fig. 1. Both materials are available all over the world and will not be depleted in centuries. The ash is helium and harmless. The repelling forces of the nuclei are overcome by auxiliary heating of a plasma, which consists of both these particle species. In the heating process particles gain energies high enough for nuclear fusion to occur and which, in turn, heats the plasma via alpha particles and delivers thermal energy via neutrons, which are both fusion products [1].

The prototype fusion reactor type under investigation is rather large with dimensions of typically 20 m in diameter and height. The stored magnetic energy reaches 40 Gigajoules. The fusion energy gain is approximately 1 GW. Unfortunately it is not possible to study the reactor processes on a smaller experimental scale. This is due to scaling laws for energy confinement and criteria for starting a self-sustaining fusion process. Typical state-of-the-art experimental set-ups are JET (Joint European Torus), ASDEX (Fig. 3) or TCV (Fig. 9), which are plasma devices with magnetic plasma confinement of the Tokamak type. In these machines the fusion processes run at a very low power level. They are physics experiments in order to understand basic questions of plasma physics with respect to nuclear fusion. In JET the goal of balancing electrical power input and fusion power output is only a factor of 8 away. With respect to this figure the progress made within the last twenty years represents more than three magnitudes, as shown in Fig. 2.

All devices constructed so far belong to the first-step machines for demonstration of the physical feasibility of nuclear fusion. Next-step machines will have to demonstrate the technical feasibility. This concerns the plasma wall interaction, thermo-mechanical stresses, fuel cycle control, ash removal and the fusion process stabilization.

Particle accelerators for nuclear physics accelerate particles of distinct kind to very high energy at very low bandwidth. In contrast, a plasma consists of a variety of particles which are in thermodynamic equilibrium, i.e. their velocity distribution is Maxwellian in all dimensions. Whereas the particles are charged the plasma in total is neutral. Ions and electrons are gyrating along the magnet field lines. Perpendicular movement is extremely hampered thus reducing radial energy transport in a toroidal device. A transformer coil is used as energy storage for inductive plasma generation and current drive. Energy exchange between particles due to Coulomb collisions prevents acceleration of electrons and causes thermodynamic equilibrium.
Fig. 1 Scheme of a fusion reactor plant

Fig. 2 Progress in fusion research
2. **THE MAGNETIC PLASMA CONFINEMENT SYSTEM**

Magnetic plasma confinement devices of the tokamak type are toroids with a strong toroidal magnet field of 3 to 5 Tesla. A high electric current in the order of Mega amperes flows in the toroidal plasma. The magnet field system consists of the toroidal field for thermal isolation and plasma confinement, and of a couple of poloidal fields for generation and stabilization of the plasma in space and shape. Most coils are energized by controlled thyristor rectifiers. Apart from the energy consuming toroidal field, which is static, all other fields are dynamic. Many of the magnet coils are magnetically coupled. Since the plasma current is inductively generated by means of a transformer coil, the plasma process in a tokamak is principally limited in time.

The engineering design of a tokamak depends on physics parameters, which in turn depend on the physics goals or even reactor requirements such as:

- energy confinement time,
- maximum plasma pressure,
- toroidal magnet field,
- plasma current,
- pulse length,
- plasma shape, radius and aspect ratio.

From these data follow technical data:

- size of the machine and magnet coils,
- magnet field energy, electric power,
- dissipated energy, thermomechanical loads.

To meet conditions which are determined by the site or the costs, a design process will yield data with respect to power and control requirements:

- normal or superconducting coils,
- minimum magnet field energy,
- mutual coupling between coils,
- transformer type: air-core or iron core,
- power supply from the mains or via a flywheel generator,
- cooling process.

Minimum magnet energy and minimum mutual coupling between coils are somewhat interdependent. The latter can be achieved by orthogonal field configuration (toroidal and poloidal, radial and vertical) or by space separation. This was the case in the ASDEX experiment illustrated in Fig. 3. A more reactor-relevant design like ASDEX-Upgrade automatically yields long-ranging magnet fields with stronger interaction.

3. **POWER SUPPLY FOR TOKAMAKS**

The choice of the power supply system depends largely on the conditions, which are given for each site by the electric energy supplier. The basic decision is then whether to use the mains directly or a flywheel generator. Energy supply from the mains means that only a few people are necessary for operation and maintenance and that
availability is a maximum. On the other hand, pulse frequency may be restricted, investment costs are raised due to harmonic frequency filters and installation of high-voltage lines and electric power stations, while operation costs may be high due to short-time and moderate energy consumption at high power. The system adopted for IPP is illustrated in Fig. 4. With flywheel generators most of the limiting factors for operation from the public power supply do not exist. The AC line frequency can be chosen to be higher for faster rectifier control though noise on the generator bus caused by a combination of differently controlled rectifiers may need special care. Energy costs are comparatively high due to low conversion efficiency, while the expenditure for qualified personnel for maintenance and operation is considerably higher.

Apart from the toroidal magnet coil all other coils need fully controlled rectifiers, some of them with current inversion. This is always true for the OH-(ohmic heating)-transformer rectifier in order to make use of twice the flux swing for plasma current build-up and sustainment at lower rated current for the coil and rectifier. It may also be useful for other circuits such as the radial field coil for vertical positioning, or for poloidal field configurations for more complex and refined tasks in plasma position and shape control.

The maximum power demand is proportional to the magnetic energy of the load:
\[ P_{\text{max}} = 2 \cdot E_{\text{magn}} \cdot (1/\tau + 1/T_F) \]

with \( \tau \) the time constant of the coil and \( T_F \) the desired minimum rise time for the current. \( \tau \) is in the range of seconds. For poloidal field coils the rise time is shorter than the time constant. An optimum matching of voltage and current ratings to the coil design will reduce the actual power demand. The maximum power easily reaches values of 100 MVA or more at currents of 50 kA.

Precise electronic control of the thyristors is a difficult task due to a rather noisy AC bus. In the case of optimum control as part of a plasma position feedback loop the requirements will be further enhanced (see section 7).

Plasma physic diagnostics are very often based upon inductive pick-up coils or detectors. The AC frequencies of the rectifier voltage output may then significantly disturb the true signal of the detector. In some cases lowpass filters in the high power DC circuit are more convenient than sophisticated correlation techniques for signal evaluation. These filters either have poor edge characteristics or show a resonance, which may interact unfavourably with a feedback loop. The design of such filters therefore requires careful investigation of high power DC and AC properties of inductors and of the feedback loop. Moreover the DC power and thermal ratings of the rectifier may rise.

Fig. 4 Modular power supply system at IPP
The current carrying plasma varies in cross section, shape, position and symmetry. It can therefore change or even disrupt on a magneto-hydrodynamic time scale in the microsecond range, and in this way induce very high voltages in loops or coils. In order to protect coils and power supplies from overvoltages, voltage limiting or switching elements are necessary [2]. The former have to be designed for high energy consumption, which is often unfeasible for technical and cost reasons, while the latter usually need additional current limiting resistors. In any case it is necessary to freewheel the DC current of the coil in case of a failure of the controlled rectifier. This can be achieved by simultaneous firing of the rectifier bridges. Short-circuiting of the rectifier transformer during operation must however, be prevented with certainty.

For current control a system of cascaded protective control loops is necessary. Power supply systems have current controllers and independent overload detecting equipment. This is especially valid for modular systems with varying loads. Overload from power loss depends on time \( t \int i^2 dt \), which is why the rectifier design is usually tailored for short-time operation in the range of some seconds. Corresponding controls for the magnet coils are independent. They are set with respect to coil parameters and include an independent current measurement. For redundancy an automatic choice of maximum current signal is recommended. Since safety factors are quite low, the protection signal levels are set depending on the desired values. Current limitation is gentle and a warning signal is given. A second protection level corresponds to the maximum tolerable values. If activated, the current is promptly switched off. In order to prevent severe plasma current disruption it is sometimes desirable to set the rectifier DC output voltage to zero instead of switching the rectifier into inverter mode.

The transformer circuit for plasma generation and plasma current build-up and sustainment is rather complex. For gas breakdown one needs a cyclic electric field of approximately 2V/m or 20V loop voltage in tokamaks of the ASDEX size. Depending on the transformer design, i.e. turns, which again depends on physical constraints such as the desired magnet flux swing, the necessary transformer voltage may exceed customary voltage levels for high current controlled rectifiers. In this case a special high voltage source has to be switched into the circuit. On ASDEX it was fortunately possible to stop using this equipment after some time of operation, since gas breakdown occurred already at 10 to 20V loop voltage, which corresponds to the rated rectifier voltage of 1 or 2 kV.

A schematic drawing of a so-called Ohmic-Heating transformer circuit is shown in Fig. 5. The transformer forms the primary winding and the plasma ring the secondary. It represents an inductive energy storage for plasma current generation and sustainment. The term Ohmic-Heating means resistive plasma heating with the plasma current. The main element is a DC breaker, which opens the current path at maximum transformer current. The current commutates into a resistor, which determines the maximum voltage at the transformer terminals and the necessary open-loop voltage for gas breakdown. The transformer current decay can be controlled by varying the resistor and by active electronic means as soon as the power decreases. The controlled rectifier for powering the transformer remains in full operation (inverter mode). It has a low source resistance so that the breaker voltage will not appear at the rectifier's terminals. After the initial build-up of the plasma current the rectifier can again fully control the transformer current and its direction, provided the rectifier can invert both voltage and current. The breaker is switched off then. During this period the transformer current derivative is feedback controlled in order to keep the plasma current constant or to follow a given programme.
Like the transformer voltage, other sources for electric current drive are used to sustain the plasma current. These sources are beams, which transfer momentum to plasma particles, or they are preferably high frequency waves which are launched into the plasma with appropriate antennas or microwave couplers. Whereas the transformer can principally drive the plasma current only for limited time, the high frequency sources can be operated continuously but at low efficiency. This enables the tokamak to be not only a pulsed machine but also a continuously operating plasma device. It now seems possible that recharging of the OH transformer could happen during HF plasma current drive in future tokamaks. Plasma parameters would vary with this cycle then. Therefore a tokamak would still be a device with the drawback of alternating physical and technical parameters.

4. **FORCES ON THE PLASMA**

4.1 **Horizontal Equilibrium**

The plasma is like an elastic and conductive torus in which the OH transformer induces a current \( J_p \). This current has a magnetic field, which is stronger in the center than outside. With increasing torus-radius \( R_p \) the magnetic energy will be reduced, so a force \( F_1 \) pushes in that direction, as shown in Fig. 6. The gas pressure \( p_p \) is constant everywhere at the plasma-surface. The difference between the outer and inner circumference leads to a second force \( F_2 \), which depends also on the torus radius \( R_p \).

A magnetic field \( B_v \) has to be applied to compensate these forces by its vector product with the plasma-current \( J_p \): \( F_3 = J_p \times B_v = F_1 + F_2 \).

The plasma can lose its gas pressure (electron density, see Fig. 7) within about 0.1 ms at a so-called disruption. The plasma ring jumps several cm to the center, whereby a voltage spike of more than 10 kV is induced and the plasma-current \( J_p \) increases due to the magnetic flux conservation [3].

The ASDEX feedback system is able to restore a 400 kA plasma back to its original position within about 8 ms (see Fig. 7) using almost the full power of a thyristor power converter (1.2 kV, 35 kA, 100 Hz-12 pulse).
Fig. 6 Forces to the plasma in horizontal direction

Fig. 7 Plasma parameters measured from a real ASDEX plasma (400 kA) controlled with thyristor converters (40 MW)
4.2 Vertical Forces

To produce a better confinement the plasma has to be elongated. A current \( J_g \) in a pair of parallel conductors above and below the plasma will do this. The produced forces (\( F_4 \)) depend on the currents \( J_p \) and \( J_g \) and their distance from each other (Fig. 8).

If the plasma current is not exactly in the center, the plasma will move in the direction of the displacement [4]. Since the plasma consists of hydrogen gas with nearly no mass the acceleration would be extremely high. The moving plasma current however induces eddy currents (\( J_i \)) in the conductive vessel giving opposite forces (\( F_5 \)) which slow the plasma movement down to controllable values.

The resulting transfer-function between plasma-motion (\( Z_p \)) and radial-field (\( B_r \)) has one zero point (NE), one nearly negligible negative root (PX) and one very critical positive root (PP).

\[
S(p) = K_{pe} \frac{p - NE}{(p - PP)(p - PX)}
\]

where \( p = \sigma + j\omega \) = complex frequency
\( K_{pe} = \) coefficient depending on configuration and plasma current
\( NE = 1/\text{vessel time-constant} \)
\( PP = \) root depending on plasma elongation and vessel conductivity

Due to the resistivity of the vessel, these currents \( J_i \) will dissipate and the moving speed can still be very high. One example is shown in Fig. 10 where a disturbance jump has been applied to the D-shaped plasma (Fig. 9) of the TCV Tokamak under construction in Lausanne [5].
4.3 \textbf{Vertical Position Control}

The control loop for the vertical position can, in a simplified way, be outlined as in Fig. 11.

\begin{figure}[h]
\centering
\includegraphics[width=\textwidth]{Fig_11.png}
\caption{Simplified block diagram of plasma-position control in a Tokamak}
\end{figure}

The open loop root loci are the following:

- \textbf{plasma:}
\begin{equation*}
\begin{array}{c}
\text{plasma:} \\
\text{px} \\
\text{ne} \\
\text{pp} \\
\end{array}
\end{equation*}

- \textbf{coil:}
\begin{equation*}
\begin{array}{c}
\text{coil:} \\
\text{pg} \\
\text{ne} \\
\text{pl} \\
\end{array}
\end{equation*}

- \textbf{vessel:}
\begin{equation*}
\begin{array}{c}
\text{vessel:} \\
\text{pg} \\
\text{pl'} \\
\end{array}
\end{equation*}

- \textbf{regulator (PID):}
\begin{equation*}
\begin{array}{c}
\text{regulator (PID):} \\
-1/tv \\
-1/tn \\
\end{array}
\end{equation*}

Knowing these values, the dominant root of the closed control loop can be calculated for each loop gain. In Fig. 12 an example is shown of the ASDEX system in a logarithmic computer plot [4].
To obtain maximum speed and minimum overshoot one has to choose the maximum loop gain where the root-locus is just leaving the range $\omega/\sigma < 1$. For that case one has 4% overshoot. This point can also be found by the Nyquist plot with the frequency $p = -\omega + j\omega$, as demonstrated in an example Fig. 13. In that diagram there are also curves for $p = -\omega + j2\omega$ which corresponds to 20% overshoot, for $p = -\omega + 4j\omega$ and for the normal frequency plot $p = j\omega$.

5. **NONLINEAR EFFECTS**

5.1 Frequency response of discrete control elements

If a sample and hold (S+H) with a clock time $T$ of a digital computer is applied, the frequency response is:

$$H(p) = \frac{1 - e^{-pT}}{p}$$

as measured in Fig.14.
Fig. 14 Nyquist plot of the sample + hold measured with FFT (fast Fourier transform)

In contrast to the S+H, thyristor static converters or choppers have no frequency dependence.

\[ G(p) = \text{const} \]

as measured in Fig. 15.

Fig. 15 Nyquist plot \( U/\alpha \) of a thyristor converter

This is only true as long as one considers the response of the same frequency. In the spectrum however there are many other frequencies (Fig. 16) which can be determined by the Bessel-functions [6] (Fig. 17).

Fig. 16 Spectrum \( U(f) \) (sinusoidally modulated) of S+H (upper curve) and thyristor (lower curve)

Fig. 17 Determination of the thyristor spectrum in relation to modulation-amplitude
5.2 Subharmonic resonances

For stability considerations of a closed loop only those frequencies which coincide with the stimulator are critical. This is the case at small amplitudes for the thyristor converter as well as for a sample and hold at the first subharmonic $f_1 = \frac{1}{2T}$ as shown in Fig. 18.

![Fig. 18 Voltage $U(t)$ of thyristor (upper curves) and S+H (lower curves) modulated with $f_1 = \frac{1}{2T}$](image)

For large amplitudes further subharmonic frequencies can be critical (Fig. 19).

- $f_n = \frac{1}{T} - n \cdot f_n$
- $f_1 = \frac{1}{2T}$
- $f_2 = \frac{1}{3T}$
- $f_3 = \frac{1}{4T}$
- $f_4 = \frac{1}{5T}$

![Fig. 19 Spectrum of thyristor voltage modulated with large amplitudes](image)

All these subharmonic frequencies have to be added to the frequency response in all phase relations [6], which leads to the circles in the Nyquist-diagram (Fig. 20).

![Fig. 20 Nyquist plot with subharmonic circles](image)

Neither the frequency response plot nor any of these different subharmonic circles should surround the point $-1$. 
In order to find the optimum regulator-parameters one can draw these diagrams for all sorts of variables. In the example of Fig. 21 this is shown for the plasma current in the ASDEX Tokamak.

Fig. 21 Determination of maximum loop gain in an ASDEX control loop, including the thyristor converter, in respect to the plasma-current

For a Tokamak with very high plasma elongation (e.g. TCV), a fast linear amplifier with a coil inside the vessel has to help the thyristor-converter to stabilize the vertical plasma position. The stability margin of the closed loop is shown in Fig. 22 depending on the growth-rate of the elongated plasma. Figures 23 and 24 show the optimum for minimum transistor power or minimum subharmonic thyristor noise.

Fig. 22 Determination of the optimum loop-gain as a function of the plasma growth rate
Normal current control loops are of the first order-delay type, so the first subharmonic frequency $1/2T$ is the most critical (see Nyquist-diagram Fig. 25 left). The oscillogram in Fig. 25 shows the step response voltage and the current in such a loop. An overshoot can be seen, which is followed by a too-low step and then a too high again etc. This demonstrates the first subharmonic oscillation [6].

Figure 26 shows the closed loop frequency response measured on the real ASDEX Tokamak [4] or [8]. This was done with a fast Fourier analyser (FFT) in combination with a function generator giving a complete set of discrete frequencies ($20, 40, 60, 80, 100, 120, 140, 160, \text{ and } 180 \text{Hz}$) to the reference input of the regulator. The system was optimized as shown in Fig. 21. The FFT-plot of Fig. 26 corresponds to the oscillogram shown in Fig. 7.

For vertical position control of a highly elongated plasma (Figs. 9, 10) a system with coils outside the vessel being controlled with thyristor converters is too slow. One needs coils more closed to the plasma driven by a fast
Fig. 26  FFT Nyquist reference diagram measured on a 400kA ASDEX plasma closed loop controlled with a thyristor converter (1.2kV, 35kA, 100Hz-12pulse). Horizontal position control corresponding to oscillogram Fig. 7.

Fig. 27  Control system for plasma-elongation and vertical position in a tokamak with optimum disturbance feedforward for elongation control and double-loop control for stabilizing the vertical position of an elongated plasma. The regulators for the thyristor converters are nonlinear, discrete and synchronized with the thyristor pulses for optimum control (see sections 6 and 7).
transistor amplifier in addition to the thyristor converter as shown in Fig. 27. The diagram Fig. 22 corresponds to this configuration.

For the elongation of the plasma an optimum regulator described in section 7. has to control the elongation current \( I_g \) very fast at the correct ratio to the plasma current as outlined in Fig. 27.

6. **Discrete Regulator for Robust Control with Thyristors**

Using linear regulators with proportional and derivative action (PD) the achieved thyristor voltage depends very much on the polarity, the amplitude and the instant of a step in relation to the phase of the thyristor (Fig. 28). Positive transient signals fire the next phase directly while negative pulses are very often ignored.

![Fig. 28 Asymmetry of the thyristor converter](image)

This dynamic rectifier effect can be reduced by increasing the T1-time of the PD regulator or by an additional low-pass filter, but this impedes a fast control. For a critical tokamak system however one has to do better. This is why we had to design a discretely operating, synchronizable, nonlinear regulator using a fast analog computer (called PDis regulator for which a patent is pending). This regulator takes into account the amplitude of the actual thyristor pulse and the derivative action time \( T_v \) by the formula:

\[
y_k = \frac{\sin \alpha}{\omega} + \frac{1}{\sigma^2 + \omega^2} \left\{ -\frac{\sigma}{\omega} \alpha \left( \sigma \cdot \cos \alpha - \omega \cdot \sin \alpha \right) - \sigma \right\}
\]

where \( y_k \): correction time-factor  
\( \alpha \): control angle (thyristor input)  
\( \sigma_N = 2\pi/p_N \): phase interval (\( p_N \) = pole number)  
\( \omega = 2\pi \): mains frequency  
\( \sigma = -1/T_v \): derivative action coefficient
The low-pass filter is replaced by an integrator, which is reset with every thyristor trigger. In this way the delay in the synchronized regulator is as short as possible and the symmetry in polarity and linearity is always an optimum (see Figs. 29 and 36).

The PDIs regulator also takes into account the available range of operation. So the derivative action is always executed without saturation-loss in a minimum delay time (see Fig. 30).

---

Fig. 29  Thyristor voltage and current controlled with PDIs regulator

Fig. 30  Derivative-action step responses of different regulators
7. OPTIMUM CONTROL WITH THYRISTOR CONVERTERS

Modern control techniques are normally based on discrete computer signals with constant sampling interval. For phase modulated switching devices the intervals vary with the amplitude (see Fig. 31). For this reason the standard algorithms cannot be applied to thyristor converters [7].

If one separates a modulated chopper or thyristor voltage from the signal, which is there without modulation, one obtains pulses of different size at varying distances (see Tab+ and Tab– Fig. 31).

![Diagram of modulated and unmodulated voltage signals of thyristor and chopper]

Fig. 31 Modulated and unmodulated voltage signals of thyristor and chopper

For discrete optimum control one can try to find a combination of pulses stimulating a step with minimum risetime and no overshoot at the output of a system $S(p)$ (see Fig. 32). The distance Tab between the first two pulses has to be a variable, while the other pulses have the average time interval $T$, as outlined in Fig. 33.
The amplitudes of the pulses $\Psi_{DA}$ and $\Psi_{B}$ have to be determined, so that the output voltage $U_\Sigma$ is stationary as soon as possible.

If the frequency response $S(p)$ is known, the time response can be calculated by the residues $R_i$.

$$S(p) = \sum_{\mu=1}^{m} \frac{N_{\mu} - p}{\pi_{\mu}} \quad \text{and} \quad R_i = -p_i \cdot \sum_{\nu=1}^{n} \frac{N_{\mu} - p_i}{\pi_{\nu}}$$

The response of the pulse $\Psi_{DA}$ is:

$$U_{DA}(t) = \Psi_{DA}(R_1 e^{P_1(t + Tab)} + R_2 e^{P_2(t + Tab)} + R_3 e^{P_3(t + Tab)})$$

and the response of $\Psi_{B}$ is:

$$U_{B}(t) = \Psi_{B}(R_1 e^{P_1 t} + R_2 e^{P_2 t} + R_3 e^{P_3 t})$$

The pulses $\Psi_{Pe}$ can be described by the average voltage $U_{Pe} = \Psi_{Pe}/T$ (see Fig. 33).

The step response $Ss$ of the voltage ($U_{Pe} = 1$) can be calculated in a similar way:

$$Ss(p) = S(p)/p$$
with the residues:
\[ R_{s0} = \frac{1}{R_s} \quad R_{s1} = \frac{R_1}{R_1} \quad -s_2 = -\frac{2}{R_2} \quad -s_3 = -\frac{3}{R_3} \quad \text{etc.} \]

The step response is now:
\[ U_p(t) = \frac{\Psi P_e}{T} \times (R_0 + R_1/p_1 \times e^{p_1t} + R_2/p_2 \times e^{p_2t} + R_3/p_3 \times e^{p_3t}) \]

All pulses together:
\[ U_\Sigma = \frac{1}{T} \frac{\Psi P_e}{T} + (\Psi_{DA} e^{p_1 T_0} - \Psi_B + \frac{\Psi P_e}{T \Omega P_1}) \times R_1 \times e^{p_1 t} \]
\[ + (\Psi_{DA} e^{p_2 T_0} - \Psi_B + \frac{\Psi P_e}{T \Omega P_2}) \times R_2 \times e^{p_2 t} \]
\[ + (\Psi_{DA} e^{p_3 T_0} - \Psi_B + \frac{\Psi P_e}{T \Omega P_3}) \times R_3 \times e^{p_3 t} \]
\[ \text{etc.} \]

If the response \( U_\Sigma \) has to be stationary as soon as possible, the coefficients of the slow elements have to be zero. This implies for the roots \( P_1 \) and \( P_2 \):

\[ (\Psi_{DA} e^{p_1 T_0} - \Psi_B + \frac{\Psi P_e}{T \Omega P_1}) = 0 \]

and

\[ (\Psi_{DA} e^{p_2 T_0} - \Psi_B + \frac{\Psi P_e}{T \Omega P_2}) = 0 \]

or:

\[ \Psi_{DA} = \frac{\Psi P_e}{T} \left( \frac{1}{P_1} - \frac{1}{P_2} \right) e^{P_2 T_0} - e^{P_1 T_0} \]

and

\[ \Psi_B = \frac{\Psi P_e}{T} \left( \frac{1}{P_1} e^{P_2 T_0} - \frac{1}{P_2} e^{P_1 T_0} \right) \]

In Fig. 34 the response to the pulses \( \Psi_{DA} \Rightarrow UDA \) and \( \Psi_B \Rightarrow UB \), as well as the step response to \( U_{Pe} \Rightarrow UP \) and the resulting voltage \( U_\Sigma \) is plotted separately. It can be seen that \( U_\Sigma \) becomes stationary after the time \(-1/P_3\)

An electronic system (Fig. 35) has been designed to make these calculations with the actual time \( T_0 \) for every pulse within about 20 \( \mu s \). An automatic saturation control similar to Fig. 30, and the nonlinear compensation mentioned in section 6 are also included. It is called a PDIsB regulator (patent pending). In Figs. 36 and 37 is shown how this regulator controls the magnetic field inside the vacuum vessel of a Tokamak.

This regulator system can operate not only in the second-order mode (PDIsB) but also in the first, like a discrete proportional and derivative action (PDIs) or as a proportional, integral and derivative (PIDis) regulator (see section 6).
Fig. 34  Example of a second-order optimum control with PDisB regulator

Fig. 35  First and second-order discrete regulator system for chopper and thyristor power converters
Fig. 36 Optimum control with thyristor converter measured on second-order system

Fig. 37 Optimum control of second-order delay system and thyristor converter with PDisB regulator and nonlinear correction

**Remark:** For optimum control with thyristor rectifiers one needs a good balanced system with a minimum of noise on the AC bus. Reproducible disturbances and even unsymmetrical phases can be equalized by computer with a synchronized transient recorder [8] (see computer feedforward Fig. 38). The rest of the errors and random disturbances have to be eliminated by the closed loop control (Fig. 38).
Fig. 38 Overview of a complete Tokamak plasma control system (see section 8)
8. CONTROL WITH PROGRAMMABLE MATRICES

The control of vertical position, as shown in Fig. 11 and Fig. 27, is only a simplified part of the Tokamak plasma control. A modern Tokamak has many control coils installed, which are used in different combinations for many tasks [9]. The distribution of the signals to the corresponding coils has to be done with a control matrix (Fig. 38).

A similar problem is to measure the position and the shape of the plasma [10]. Many different measuring signals are needed to determine one of the control parameters. For these calculations one needs another matrix (see measuring matrix Fig. 38).

The matrix calculations have to be carried out very rapidly, so that for every thyristor pulse the actual signal is available without delay, since digital computers and even modern transputers are still too slow. Therefore we have designed a special CAMAC module (called AMUVECAM see Fig. 39) for vector and matrix calculations in hybrid technique [11]. The calculations are made using analog components, which are programmed digitally. The multiply bandwidth is about 30 kHz and all coefficients can be completely reprogrammed within 4ms from a local memory (12 x 8192 x 16bit) in clocked mode or with random access.

![CAMAC module for vector or matrix multiplication (AMUVECAM)](image)

With this module all sorts of preprogrammed calculations can be made for plasma shape and position control as well as for simple programming of the coil currents or the magnetic field inside the vessel (via the PDisB regulator).

The vessel, the coils and the supporting structure of a Tokamak have a complicated dynamic behaviour, since they are all magnetically coupled. It would be too much work to calculate numerically all these time dependent eddy currents and fields and the corresponding coefficients for the control matrix. Another easier way is to determine the matrix coefficients experimentally at the real Tokamak. One can open the control-loop at the control-parameters $\Psi$ (see Fig. 38) and give a voltage step to one input of the "PI / D" regulator (integrator off).
By varying the digitally programmable coefficients of the matrix and the regulators one has to bring the corresponding output of the measuring matrix to a maximum while all the other signals must remain as close to zero as possible.

The multiplying components of AMUCECAM are fast analog devices, so it is possible to produce feedback from the output of the matrix to an input (see Fig. 40). In some cases this is easier for dynamic compensation than the forward control.

Fig. 40 Control matrix including program generator, decode matrix and different parts of a dynamic decouple-matrix

The coils are at different distances from the vessel and the plasma, which leads to different delays for the magnetic field responses. This can be equalized by the coefficients of the PDis regulators and by supplements to the coil resistors (see Fig. 38).

9. **CONCLUSIONS**

Specially designed discrete regulators can improve the control with choppers or thyristor converters. A computer program, taking into account the subharmonic frequencies, finds the best regulator for robust or optimum control. For multi-parameter systems a fast programmable matrix calculator is effective for many
different purposes such as program control, dynamic decoupling, disturbance feedforward and closed loop control with fast switching power converters.

Control problems of a Tokamak lead to technical solutions which may also be useful for other applications where high power and very fast control are required.

* * * *

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BATTERY ENERGY STORAGE FOR ELECTRIC VEHICLES

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Abstract

The advanced battery system NaS is described. Sodium and sulphur are the liquid electrodes and β-Al₂O₃-ceramic is used as a solid electrolyte. The gravimetric energy density of batteries based on this system is four times higher than the energy density of conventional lead acid batteries. Design characteristics and performance of electric vehicles based on this battery system are described.

1. Introduction

The utilization of electric energy is widely spread in our industrialized community inspite of its dependence on wire connections between a power plant and the end user. Such a system would not have survived in nature during evolution because of its vulnerability to faults in the distribution network. This drawback has been mastered by the utilities by establishing the European linked network and centralized power plants. If an appropriate storage facility for electric energy is available, the distribution network can develop to a more decentralized interlinked structure and even most sorts of automobiles can be electrified. This of course depends on the viability of a battery system that can generally be used for the storage of electric energy.

The following years will show, whether the electrochemical system sodium and sulphur with β-Al₂O₃-ceramic as a solid electrolyte as described below is this system.

2. The principle of the battery

The energy turn-over \( \Delta H \) of all chemical reactions is described by the Gibbs-Helmholz-equation:

\[
\Delta H = \Delta G + T \, dS
\]

with the change of free energy \( \Delta G \) and the inner energy \( T \, dS \).
In batteries the electro-chemical reaction is controlled by the electrolyte called separator in order to utilize the free energy.

For a technical battery system with a high energy density the following conditions have to be met:

- The electrodes should be material from the first group in the periodic system of elements (donator) and the fifth or sixth group (acceptor)
- A separator compatible with the electrode materials must be available in order to control the reaction
- All materials must be available for low costs
- The reaction must be reversible for an accumulator
- The system must be technically feasible and safe.

Many systems have been investigated in the last decades but the system with sodium and sulphur as electrodes and \( \beta\)-\( \text{Al}_2\text{O}_3 \) ceramic as electrolyte seems to be a favorite.

The basic reaction is:

\[
2 \text{Na} + 3\text{S} \rightleftharpoons \text{Na}_2\text{S}_3.
\]

The atomic weight of \( \text{Na}_2\text{S}_3 \) is 142.2 g. One mol produces a charge of 53.3 Ah at an electrochemical potential of 2.1 V. This results in a theoretical energy density of 790 Wh/kg. Comparable figures for the well known systems \( \text{Pb}/\text{H}_2\text{SO}_4/\text{PbO} \) is 161 Wh/kg with 1.93 V and \( \text{Cd}/\text{KOH}/\text{NiOOH} \) is 208 Wh/kg with 1.24 V.

3. **How the battery works**

   With sodium and sulphur as electrodes and \( \beta\)-\( \text{Al}_2\text{O}_3 \) ceramic as separator the operating temperature has to be 300°C for the electrodes to be liquid [1, 2, 3, 4].

The principle of the sodium-sulphur cell is shown in Figure 1. The sodium and sulphur reactants, each in liquid form in enclosed containers, are separated from each other by the ceramic electrolyte, which conducts sodium ions and acts as an insulator for electrons. This property is possessed, for example, by a material consisting of aluminium oxide, sodium oxide, and magnesium oxide.
The electronic current flowing through the external load resistor during discharge corresponds to a flow of sodium ions through the electrolyte from the sodium side to the sulphur side, where sodium polysulphide is formed. The voltage assigned to the chemical reaction is between 2.08 and 1.78 V, depending on the degree of discharge involved (Fig. 2)

Figure 2
Open circuit voltage of a sodium sulphur cell, and voltage during charging and discharging at constant current, plotted against the degree of charge
In order to avoid solidification polysulphide (this would involve a significant increase in the internal resistance), a minimum operating temperature of 285 °C is required. The end of the charging process is characterized by a clear rise in the internal resistance of the cell, thus enabling this state to be easily detected. There are no other chemical reactions involved like the gasing familiar from lead batteries, so that the coulometric efficiency is always 100%.

Actual sodium sulphur cells are designed in cylindrical form, as shown in Fig. 3. A technical cell contains additional elements, comprising a carbon felt in the sulphur compartment, enabling the electrons to be discharged to the housing through the electrically insulating sulphur. An electrode is installed in the sodium compartment, enabling the current flow through the external circuit even if the sodium level is dropping. The electrode is located in close proximity to the electrolyte tube, ensuring that due to capillary force the entire internal surface of the electrolyte is wetted with sodium. On the other hand, in the event of a forced fracture of the electrolyte only a small amount of sodium will come into direct contact with the sulphur, so that the exothermic reaction results in only a slight increase in temperature. The current collector thus simultaneously performs a safety function.

![Figure 3](image-url)
A cell with a capacity of 35 Ah has a diameter of 35 mm and a length of 230 mm. Its internal resistance is 10 mΩ. 144 cells of this type can be installed in a battery housing with a volume of 75 dm$^3$. The housing serves as thermal insulation. Besides the electrically insulating cell mountings, the case also contains the cooling and the equipment for heating the battery and maintaining it at the required temperature. Figure 4 shows a development sample with external dimensions of 733 mm x 318 mm x 334 mm. The terminals on top of the battery case are for current connection bushings, and for connecting battery heat and temperature probes. The flange in the centre is required for evacuating the thermal insulation during manufacture. The cells account for 58% of the overall weight of 98 kg.

Figure 4
High Energy Battery Type B15
When the cells are electrically connected in four parallel strings of 36 cells each, the battery has an open circuit voltage of 75 V to 64 V. In this case the internal resistance measures 90 mΩ, and the capacity is 140 Ah. The voltage drop under load is a product of internal resistance and battery current, with the internal resistance remaining constant for all degrees of charge, both during charging and discharging.

The electrical energy which can be drawn from the battery is shown in Fig. 5 as a function of the (constant) discharge power. With a complete discharge in two hours, the energy content is e.g. 32 kWh, corresponding to an energy density of around 100 Wh/kg. The associated discharge efficiency is 92 %. Owing to the factor of charge being 1, the charging efficiency is almost exactly the same for a two-hour charging time, at approx. 91 %. If charging is spread over ten hours, which is a typical period for an electric car storage system, the charging efficiency is 98 %. Complete discharge at constant power without interruption is possible in a minimum of one hour, an 80 % discharge in less than three-quarters of an hour. This maximum continous load is determined by the dimensioning of the battery cooling system. Thanks to the high operating temperature, the cooling air flow required is only 1 m³/min.

Figure 5
Energy content as a function of the constant power drawn, for depth of discharge 0.8 and 1
That portion of the joulean heat losses not removed by the cooling system is stored in the heated-up cells, and covers the heat losses for up to 30 h. During longer standstill periods, however, additional heating must be supplied. The energy required for this can be obtained either from the electrical mains (e.g. from a power socket during waiting periods) or from the battery itself.

Thanks to the effective thermal insulation [5] the power losses of the battery are only 80 W, so that when fully charged it can maintain its own temperature for a period of some days. The vacuum-type thermal insulation is designed to withstand atmospheric pressure, and has a thermal conductivity value of 0.0025 W/mK, which is thus about 20 times lower than for the mineral-wool insulation in common use.

In order to maintain the battery in a state of operational readiness, its temperature should not be allowed to fall below the minimum operating level. It takes about 4 days to heat up the battery from a cold state, using the built-in heating conductor. It is, however, inadvisable to exceed a total of 30 freeze-thaw cycles.

The disadvantage of heat losses resulting from the high operating temperature is balanced by the advantage that this battery works independently of the ambient temperature. It is immaterial whether it is operated in an ambient temperature of -50 °C or of 80 °C.

4. Safety aspects

From the very beginning of the development safety aspects have been considered:

For road vehicle applications, the batteries must be vibration resistant and accident proof. Vibration tests to DIN VDE 0122 [6] at 10 to 150 Hz, 1 octave/min, 20 m/s², and a ten-minute test at resonant frequency and 20 m/s² are passed without difficulty. The battery remains undamaged in crash tests at 50 km/h, braked in 0.08 s.

When the battery housing is deformed in a compression device with a ram 80 mm in diameter, no reactands escape. This requires a deforming force of 180 kN. After 5 minutes continuous exposure to fire (50 l of petrol completely burned underneath the battery), the interior of the battery exhibited a temperature rise of only 5 K.
5. **Properties**

The only "maintenance" required is to keep the battery at the correct temperature, a task which is performed fully automatically by a battery management unit. In contrast to conventional batteries, the system requires no topping up with water. This, plus the high energy and power densities described above, together with the high charging and discharging efficiencies, are characteristics excellently suited to the electric car application.

Two types of test batteries (10 kWh and 22 kWh) are presently produced in a pilot plant that is run with a capacity of 3 - 10 MWh per year (Table 1 and 2). These batteries are available for vehicles and other applications.

### Table 1
**Performance data of test battery type B11**

<table>
<thead>
<tr>
<th>Dimension</th>
<th>Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>L x W x H</td>
<td>1420 x 485 x 360</td>
<td>mm</td>
</tr>
<tr>
<td>Energy (2 h)</td>
<td>22</td>
<td>kWh</td>
</tr>
<tr>
<td>Energy density (2 h)</td>
<td>88.7 (79.7)</td>
<td>Wh/l (Wh/kg)</td>
</tr>
<tr>
<td>Weight</td>
<td>276</td>
<td>kg</td>
</tr>
<tr>
<td>Thermal losses</td>
<td>ca. 170</td>
<td>W</td>
</tr>
<tr>
<td>Rated Voltage</td>
<td>90</td>
<td>120-180</td>
</tr>
<tr>
<td>OCV (100-% charge)</td>
<td>93.6</td>
<td>124.8</td>
</tr>
<tr>
<td>Minimum discharge OCV</td>
<td>80.1</td>
<td>105.8</td>
</tr>
<tr>
<td>Parallel paths</td>
<td>8</td>
<td>6</td>
</tr>
<tr>
<td>Internal resistance (310 °C)</td>
<td>56</td>
<td>100</td>
</tr>
<tr>
<td>Capacity</td>
<td>280</td>
<td>210</td>
</tr>
<tr>
<td>Peak current (&lt;3 min)</td>
<td>350</td>
<td>300</td>
</tr>
<tr>
<td>Continuous discharge current</td>
<td>280</td>
<td>210</td>
</tr>
<tr>
<td>Continuous charging</td>
<td>140</td>
<td>105</td>
</tr>
<tr>
<td>Terminals</td>
<td>front side, 60 mm additional length</td>
<td></td>
</tr>
</tbody>
</table>

### Table 2
**Performance data of test battery type B15**

<table>
<thead>
<tr>
<th>Dimension</th>
<th>Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>L x W x H</td>
<td>733 x 318 x 334</td>
<td>mm</td>
</tr>
<tr>
<td>Energy (5 h)</td>
<td>10</td>
<td>kWh</td>
</tr>
<tr>
<td>Energy density (5 h)</td>
<td>128 (102)</td>
<td>Wh/l (Wh/kg)</td>
</tr>
<tr>
<td>Weight</td>
<td>98</td>
<td>kg</td>
</tr>
<tr>
<td>Thermal losses</td>
<td>ca. 80</td>
<td>W</td>
</tr>
<tr>
<td>Rated Voltage</td>
<td>24</td>
<td>36</td>
</tr>
<tr>
<td>OCV (100-% charge)</td>
<td>24.9</td>
<td>37.4</td>
</tr>
<tr>
<td>Minimum discharge OCV</td>
<td>21.4</td>
<td>31.1</td>
</tr>
<tr>
<td>Parallel paths</td>
<td>12</td>
<td>8</td>
</tr>
<tr>
<td>Int. resistance (310 °C)</td>
<td>10</td>
<td>22.5</td>
</tr>
<tr>
<td>Capacity</td>
<td>420</td>
<td>280</td>
</tr>
<tr>
<td>Peak current (&lt;3 min)</td>
<td>200</td>
<td>200</td>
</tr>
<tr>
<td>Current up to thermal limit</td>
<td>200</td>
<td>200</td>
</tr>
<tr>
<td>Continuous discharge current</td>
<td>150</td>
<td>100</td>
</tr>
<tr>
<td>Continuous charging</td>
<td>45</td>
<td>30</td>
</tr>
<tr>
<td>Terminals</td>
<td>top side, local 27 mm additional length</td>
<td></td>
</tr>
</tbody>
</table>
The expected life duration in operation is more than one year or 600 full cycles corresponding to e.g. 60,000 km whereas the shelf life is unlimited if stored at ambient temperature.

Figure 6 shows that the practical advantages (maintenance free, high efficiency, low cost potential, recycling, independence on ambient temperature) are combined with high energy density at high power by a factor of four compared to lead acid batteries.

Figure 6
Battery Properties

This as well as the ability for rapid cycles (1 h discharge, 2 h charge) recommend this High Energy Battery to power electric vehicles to common use. Test vehicles equipped with this type of battery achieved already typical performance values as given in Table 3.
Table 3
Performance data of test vehicle BMW 3 series

<table>
<thead>
<tr>
<th>Status:</th>
<th>Test vehicle (not commercially available)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Manufacturer:</td>
<td>BMW Germany</td>
</tr>
<tr>
<td>Base vehicle:</td>
<td>BMW 325 ix</td>
</tr>
<tr>
<td>Seats:</td>
<td>4</td>
</tr>
<tr>
<td>Kerb weight:</td>
<td>1250 kg (incl. battery)</td>
</tr>
<tr>
<td>Gross vehicle weight:</td>
<td>1735 kg</td>
</tr>
<tr>
<td>Dimensions:</td>
<td>L = 4,33 m W = 1,66 m H = 1,40 m</td>
</tr>
<tr>
<td>Battery:</td>
<td>ABB High Energy Battery</td>
</tr>
<tr>
<td>(System Sodium Sulfur = NaS)</td>
<td></td>
</tr>
<tr>
<td>Motor:</td>
<td>ABB, DC-motor, separately excited</td>
</tr>
<tr>
<td>Transmission:</td>
<td>20 kW (continuous)</td>
</tr>
<tr>
<td>On board charger:</td>
<td>Two speed manually shifted</td>
</tr>
<tr>
<td>Heating:</td>
<td>ABB, 220 V, 2 kW</td>
</tr>
<tr>
<td>Test weight:</td>
<td>Webasto (fuel)</td>
</tr>
<tr>
<td>c_A:</td>
<td>1350 kg</td>
</tr>
<tr>
<td>Max. speed:</td>
<td>0,67 m²</td>
</tr>
<tr>
<td>Range at 50 km/h:</td>
<td>Intentionally limited to 95 km/h</td>
</tr>
<tr>
<td>Range at 90 km/h:</td>
<td>170 km</td>
</tr>
<tr>
<td>Range at ECE city cycle:</td>
<td>135 km</td>
</tr>
<tr>
<td>Acceleration 0...50 km/h:</td>
<td>150 km</td>
</tr>
<tr>
<td>Consumption at AC-outlet:</td>
<td>9 s</td>
</tr>
<tr>
<td>Charging time:</td>
<td>31 kWh/100 km (practical driving including battery heating energy)</td>
</tr>
<tr>
<td>overnight (14/14 km)</td>
<td></td>
</tr>
</tbody>
</table>

References:


A SIMULATION MODEL FOR SHAFT GENERATOR PLANTS

J.-P. Jensen,

Abstract

A simulation model to describe the dynamic behaviour of shaft generator plants with frequency converter is introduced. The differential equations for the dynamic behaviour are calculated by a numeric integration. The components of the plant are described in the two-axis transformation. A converter with an AC mains is reduced to a DC equivalent circuit with temporal parameters. The inductive circuits are summarized in a way to calculate the voltage explicitly. This means an enormous reduction of calculation time. So personal computers can be used with acceptable calculation time and accuracy.

1 Shaft Generator

Shaft generators have been used on motorships for 30 years. They are driven by the main engine and produce the electric energy instead of the auxiliary engines. The consumption of the mains is less than 10% of the propulsion power.

Figure 1 gives an overview of the machine room of a merchant ship. The main engine drives the propeller and is usually running with heavy and cheap fuel. The electricity is produced by the auxiliary machines. These engines need marine fuel which is twice as expensive. Due to the fact that the shaft generator is driven by the main engine, cheap fuel is used and the running time of the auxiliaries is reduced. If a directly coupled shaft generator is used, the frequency is proportional to the r.p.m. of the shaft. To achieve a stable frequency, a frequency converter shown in Fig. 3 is necessary. The frequency converter consists of a rectifier, an inverter and a synchronous condenser. The 3-phase rectifier converts the AC current into DC current. The inverter is externally commutated and needs reactive power because of the trigger delay angle. The synchronous condenser delivers the reactive power for the inverter, for the consumers and also the short circuit current to blow the fuses. It determines the stability of the mains. The pony motor starts the synchronous condenser.

The inverter delivers the active power and the synchronous condenser the reactive power. The rated power of the synchronous condenser is nearly the same as the rated power of the mains, which is between 500 kVA and 5 MVA.

The reason for the use of an externally commutated inverter is the price. This type of inverter is much cheaper and more reliable than a self commutated inverter.
2 Why simulations and not measurements?

Thyristor-controlled shaft generator plants are expensive, they cost more than DM 500.000. The electric power is most important for the safety of a ship. Measurements for modifications or short circuit studies might damage the electrical equipment. This can cause distress on sea.

The aim of my study was to reduce costs and to increase the efficiency of the shaft generator plant. This can be done by decreasing the synchronous condenser and increasing the capacitors of the harmonic filter. But a small synchronous condenser causes less stability of the mains and therefore less reliability and safety of the ship.

In this report I will point out a method for simulating thyristor-controlled rectifiers. Particle accelerators have these rectifiers in great number.

This simulation model neglects the small time constants and gets rid of the enormous spread of the time constants in rectifiers. This makes the simulation very fast and allows the use of a desk top or personal computer.

3 Dynamic Model of the Three-Phase Equipment

3.1 Two-axis Theory

For describing the dynamic behaviour it is advantageous to use the two-axis theory of synchronous machines (R.H. Park, 1929). The two-axis theory is also possible for three-phase chokes and three-phase capacitors. The equations are shown in the next chapter.

There are two pointers in the two-axis transformation, one in the direct axis and the other in the quadrature axis. The d-pointer shows to the pole shoe and the q-pointer is right-angled to the pole shoe. The d-pointer and the q-pointer are rotating with the rotor. Under steady state conditions, the d and q pointers are DC values. The three-phase values are alternating. The equations for the transformation from the two-reaction theory to the three-phase model are

\[
\begin{pmatrix}
  u_1 \\
  u_2 \\
  u_3 
\end{pmatrix}
= \begin{pmatrix}
  \cos \gamma - \sin \gamma \\
  \cos(\gamma - 2\pi/3) - \sin(\gamma - 2\pi/3) \\
  \cos(\gamma - 4\pi/3) - \sin(\gamma - 4\pi/3)
\end{pmatrix}
\begin{pmatrix}
  u_d \\
  u_q
\end{pmatrix}
\]

\[
\begin{pmatrix}
  i_1 \\
  i_2 \\
  i_3
\end{pmatrix}
= \begin{pmatrix}
  \cos \gamma - \sin \gamma \\
  \cos(\gamma - 2\pi/3) - \sin(\gamma - 2\pi/3) \\
  \cos(\gamma - 4\pi/3) - \sin(\gamma - 4\pi/3)
\end{pmatrix}
\begin{pmatrix}
  i_d \\
  i_q
\end{pmatrix}
\]

\(u_1, u_2, u_3, i_1, i_2, i_3,\) three-phase voltage and current

\(u_d, u_q, i_d, i_q\) two-axis transformation voltage and current

\(\gamma = \int \omega \cdot dt\) load angle
3.2 Synchronous Condenser

The synchronous condenser is a synchronous machine with damper windings. Fig. 4 shows this machine in the 2-axis transformation.

\[ u_d = R_d \cdot i_d + d(\psi_d)/dt + e_d \]  
(1)  
\[ u_q = R_q \cdot i_q + d(\psi_q)/dt + e_q \]  
(2)  
\[ \sigma = R_D \cdot i_D + d(\psi_D)/dt \]  
(3)  
\[ \sigma = R_Q \cdot i_Q + d(\psi_Q)/dt \]  
(4)  
\[ u_f = R_f \cdot i_f + d(\psi_f)/dt \]  
(5)

with the induced voltages

\[ e_d = -\omega \cdot \psi_q \]  
(6)  
\[ e_q = \omega \cdot \psi_d \]  
(7)

The shaft generator is also a synchronous machine but without damper windings just to reduce the current harmonics. In this case only the equation for the stator and rotor excitation are needed.

3.3 Three-phase Choke

It is also possible to find the equations for a three-phase choke in the two-axis transformation in Fig. 5.

\[ u_d = R_d \cdot i_d + L_d \cdot d(i_d)/dt + e_d \]  
(8)  
\[ u_q = R_d \cdot i_q + L_d \cdot d(i_q)/dt + e_q \]  
(9)  
\[ e_d = -\omega \cdot L_d \cdot i_q \]  
(10)  
\[ e_q = \omega \cdot L_d \cdot i_d \]  
(11)

The two-axis transformation induces the additional voltages \( e_d \) and \( e_q \).

3.4 Three-phase Capacitor

Figure 6 shows the three-phase capacitor in the two-axis transformation.

The equations are:

\[ i_{dc} = C \cdot d(u_{dc})/dt + i_{d, rot} \]  
(12)  
\[ i_{qc} = C \cdot d(u_{qc})/dt + i_{q, rot} \]  
(13)
\[ i_{\text{rot}} = -\omega \cdot C \cdot u_{\text{ac}} \quad (14) \]
\[ i_{\text{grot}} = \omega \cdot C \cdot u_{\text{dc}} \quad (15) \]

For a three-phase capacitor the two-axis transformation induces the additional currents \( i_{\text{rot}} \) and \( i_{\text{grot}} \).

### 3.5 Three-phase Six-pulse Bridge

The three-phase AC mains and the DC intermediate circuit are coupled by a six-pulse thyristor bridge. For the simulation of the dynamic behaviour of the shaft generator plant it is sufficient to modulate the thyristors as ideal switches. The thyristor bridge forces an asymmetrical operation of the AC mains.

The aim is to find the equations for coupling the AC and DC circuits. The AC side of the six-pulse bridge is determined by the two voltages \( u_p \) and \( u_q \) and by the two currents \( i_p \) and \( i_q \). The allocation is given in Fig. 7 and Fig. 8. During the non-commutation cycle the commutation current \( i_s \) is zero and during the commutation cycle the commutation voltage \( u_k \) is zero.

The equations of the two-axis transformation for a six-pulse bridge are

**Non-commutation:**

\[
\begin{pmatrix}
 u_d \\
 u_q \\
 i_d \\
 i_q \\
\end{pmatrix} = \frac{2}{3} \begin{pmatrix}
 \cos(\gamma + \alpha - 30^\circ) & \sin(\gamma + \alpha) \\
 -\sin(\gamma + \alpha - 30^\circ) & \cos(\gamma + \alpha) \\
\end{pmatrix} \begin{pmatrix}
 u_p \\
 u_q \\
 i_p \\
 i_q \\
\end{pmatrix}
\]

<table>
<thead>
<tr>
<th>Cycle</th>
<th>1 N</th>
<th>2 N</th>
<th>3 N</th>
<th>4 N</th>
<th>5 N</th>
<th>6 N</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \alpha )</td>
<td>30°</td>
<td>330°</td>
<td>270°</td>
<td>210°</td>
<td>150°</td>
<td>90°</td>
</tr>
</tbody>
</table>

**Commutation:**

\[
\begin{pmatrix}
 u_d \\
 u_q \\
 i_d \\
 i_q \\
\end{pmatrix} = \frac{2}{3} \begin{pmatrix}
 \cos(\gamma + \alpha) & \sigma \\
 -\sin(\gamma + \alpha) & \sigma \\
\end{pmatrix} \begin{pmatrix}
 u_p \\
 u_q \\
 i_p \\
 i_q \\
\end{pmatrix}
\]

<table>
<thead>
<tr>
<th>Cycle</th>
<th>1 K</th>
<th>2 K</th>
<th>3 K</th>
<th>4 K</th>
<th>5 K</th>
<th>6 K</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \alpha )</td>
<td>0°</td>
<td>300°</td>
<td>240°</td>
<td>180°</td>
<td>120°</td>
<td>60°</td>
</tr>
</tbody>
</table>

### 3.6 Conduction through of a six-pulse bridge

After a missing commutation the firing of a next thyristor pair causes a conduction through of one bridge leg. A conduction through means a short circuit in the DC circuit. The three possible types of conduction
through are shown in Fig. 9.

The conducting through of a bridge can be split into a normal operating bridge and a short circuit in the DC intermediate circuit. The equations of chapter 3.5 are still valid.

4 Coupling of Electrical Components

4.1 Equivalent circuit for a synchronous machine with a three-phase bridge

The three-phase bridge couples the synchronous machine to the DC circuit. The thyristors are assumed as ideal switches. During the non-commutation time, the DC circuit is connected to two phases of the synchronous machine. During the commutation time the DC circuit is connected to three phases but two of them are shortened. The terminal characteristic of the DC circuit and the connected AC lines are inductive. Therefore the equivalent circuit is a resistor, a choke and a voltage source as shown in Fig. 10 and Fig. 11.

\[
di_p/dt = (u_i - u_{cc} - R_{cc} \cdot i_p)/L_{cc}
\]

(16)

The general equation for an inductive circuit is

\[
di_p/dt = A \cdot u_i + H
\]

(17)

The detail equations can be read in the author's thesis.

4.2 Equivalent circuit for parallel circuits

A lot of components are in parallel to the AC mains. A simple rule to find an equivalent inductive circuit for parallel inductive circuits is shown below. Each circuit is given by the equation

\[
di_v/dt = A_v \cdot u_v - H_v.
\]

(18)

Kirchhoff's law for a knot is

\[
i = \sum_v i_v
\]

(19)

\[
di/dt = \sum_v di_v/dt
\]

(20)

\[
u = u_v
\]

(21)

The equivalent circuit has the equation

\[
di/dt = A \cdot u + H
\]

(22)

The values for \( A \) and \( H \) are

\[
A = \sum_v A_v
\]

(23)
\[ H = \sum \nu H_n \]  

(24)

4.3 Equivalent circuit for series circuits

The rule is similar to the one in chapter 4.2. The Kirchhoff's law for series circuits is

\[ u = \sum \nu u_n \]  

(25)

\[ i = i_n \]  

(26)

Each inductive circuit is given by

\[ u_n = \frac{1}{A_n} \cdot \frac{di_n}{dt} - \frac{H_n}{A_n} \]  

(27)

With Kirchhoff's law we get

\[ u = \sum \nu \frac{1}{A_n} \cdot \frac{di}{dt} - \sum \nu \frac{H_n}{A_n} \]  

(28)

The differential quotient is

\[ \frac{di}{dt} = \frac{1}{\sum \nu \frac{1}{A_n}} \left( u + \sum \nu \frac{H_n}{A_n} \right) \]  

(29)

The values for the equivalent circuit are

\[ A = \frac{1}{\sum \nu \frac{1}{A_n}} \]  

(30)

\[ H = A \cdot \sum \nu \frac{H_n}{A_n} \]  

(31)

5 Calculation of the Shaft Generator Plant with Rectifier and Inverter

The shaft generator plant and the mains are placed together and the extensions are only some ten to hundred meters away. There are no transmission line affects and therefore the equations are linear inhomogeneous differential equations. They are first order or can be reduced to single differential equations of first order. The temporal curve can be calculated by numeric integration step by step. The numeric integration is done by EDPCs or personal computers. At the beginning of the 80s there were no converter simulation programs for personal computers available. This was the reason to look for a very fast algorithm. The point for a fast algorithm is the smallest possible spread of time constants and the smallest possible inversions of matrixes.

There are several numeric integration methods known. The author uses an integration with step control, the so-called Merson integration.
5.1 Terminal Characteristics of Electrical Circuits

The components of a shaft generator plant are not coupled without feedbacks. They exchange energy at the terminals. Therefore the terminal characteristics of electrical circuits must be taken into consideration. The classification of terminal characteristics is inductive, capacitive and resistive. The equations have state variables and disturbance variables. A state variable changes its values continually and a disturbance variable can have any temporal course, even steps. If an inductive and a capacitive circuit are coupled, the state variable of one circuit is the disturbance variable of the other circuit. For a resistive circuit the disturbance variable can be easily calculated through the state variable of the other circuit. The inhomogeneous part is explicitly known and the calculation causes no problem.

If one couples the circuits with equal characteristic, the disturbance variable is not known. The variable is implicit and several solutions are known. For example cable capacitors or bleeder resistors can be added or the disturbance variables can be calculated by iteration (Newton procedure). The disadvantage is an enormous increase of integration steps which means more calculation time.

For the shaft generator plant the disturbance value of the DC link between rectifier and inverter is the voltage (see Fig. 12). It can easily be calculated explicitly. The rectifier and the inverter circuits are inductive equivalent circuits (see Fig. 13). The general equations are

\[
\frac{d i_{re}}{dt} = A_{re} \cdot u_{re} + H_{re} \quad \text{rectifier} \\
\frac{d i_{in}}{dt} = A_{in} \cdot u_{in} + H_{in} \quad \text{inverter}
\]

Kirchhoff's law says

\[
\frac{d i_{re}}{dt} + \frac{d i_{in}}{dt} = 0
\]  \hspace{1cm} (34)

\[
u_{re} = u_{in}
\]  \hspace{1cm} (35)

Equations (34) and (35) make

\[
A_{re} \cdot u_{re} + H_{re} + A_{in} \cdot u_{re} + H_{in} = 0
\]

\[
u_{re} = \frac{H_{re} + H_{in}}{A_{re} + A_{in}}
\]  \hspace{1cm} (37)
For \( n \) parallel circuits the terminal voltage \( u \) is explicitly
\[
  u = \frac{\sum E_n}{\sum A_n}
\]  \hspace{1cm} (38)

5.2 Calculation of the Switching Time

The thyristors and the diodes are switched on or off at appointed times. These times are not related to the integration steps and it may happen that the switching time is somewhere within the interval. This mismatch means that the voltage curves are wrong and also contain wrong subharmonics. The inverter can conduct through because of incorrect delay triggers. The switching time can be calculated as shown below. Figure 14 shows the switching off of the current through the thyristor. After the calculation of the integration step, the value for the current of the thyristor is negative. A negative current is not possible for an ideal thyristor because it will switch off during the zero crossing. The switching time can be calculated by the "regular falsi". The time difference to the zero crossing is
\[
  T_{\text{switch}} = D t \cdot \frac{i}{i - i_0}
\]  \hspace{1cm} (39)

5.3 Simulation Program

The simulation program is written in HP Basic 3.0 for a Hewlett Packard desktop computer. The program is oriented to equations and not, like other simulation programs, to the electrical components.

The principle of calculation is described as follows. First of all, the program has to calculate the disturbance variables of the equations. The state variables and the parameters are known. The three-phase systems are converted to the two-axis transformation. The equations of the ship mains are summarized to an equivalent synchronous machine as shown in chapter 4.2 and 4.3. The ship mains consist of a synchronous condenser, a choke, a harmonic filter, an asynchronous motor and an inductive load. The frequency is variable.

The equivalent synchronous machine is coupled through the inverter and the rectifier to the shaft generator. The coupling equations of the two-axis transformation for a six-pulse bridge are given in chapter 3.5 and 3.6. At last the voltage of the DC intermediate circuit can be calculated (see chapter 5.1). At this point the differential quotient of the intermediate current can be calculated as well as all the other values. The calculation goes from the intermediate circuit to the shaft generator and to the ship mains. Finally the back transformation to the three-phase notation can be done. This can be necessary if one wants to plot the curves of the real three-phase AC system.

Figure 15 shows the flow chart of the simulation program. The numeric integration is the Merson algorithm with step control but there are other algorithms possible. The integration subroutine uses the principle of calculation as mentioned above.

The equations depend on the rectifier status and the inverter status. The status is either commutation.
non-commutation or conduction through. After one integration step the subroutine "gate control set" checks if the current status of the rectifier and inverter are valid. If not, the switching time is calculated and the last integration is repeated with the new time step. Then the new cycle is selected and the converter equations are changed. Finally, the values are stored and plotted.

5.4 Plots of the Simulation

At the end I will show some plots. The parameters are related to the rated impedance of the plant. The rated Y impedance has the value 1. The rated voltages and the rated currents also have the value 1. The time scale is seconds.

The first set of plots shows the current and voltage curves of a synchronous generator with six-pulse bridge and an inductive DC circuit. The first plot shows the three-phase curves and the second the corresponding voltages \( u^c \) and \( u^q \) of the two-axis transformation. The commutation notches of the bridge can also be seen in the transformed voltage curves of the stator. The third plot shows the transformed currents of the stator and the currents of the damper windings \( i_D \) and \( i_Q \). The six-pulse bridge always forces current in the damper windings, even under steady state conditions. This means additional losses on the stator and has to be taken into account.

The simulation of a shaft generator plant from the AEG company is shown in the fourth plot. The synchronous condenser (Phasenschieber) is a normal 4-pole synchronous machine for 60 Hz. The shaft generator is a 16-pole slow-turning (100 r.p.m.) synchronous machine without damper windings. The generator is directly mounted to the B-side of the main engine. The plot shows a long commutation period and a small non-commutation period of the shaft generator current \( I_{sg} \). The voltage curve of the DC circuit can be seen on the generator voltage \( U_{12_{sg}} \). The voltage curve of the synchronous condenser \( U_{12_{sc}} \) has the commutation notches of the inverter. The current \( I_{12_{sc}} \) is not sinusoidal and causes additional losses. This means a reduced efficiency of the plant and this is a small disadvantage in comparison to other types of shaft generator plants.

6 Conclusion

A simulation model for a frequency converter between two stand alone AC systems is introduced. The frequency of the generator alternates with the turning of the main shipmachine. The generator current is converted into DC by a six-pulse diode or thyristor bridge. The DC is inverted into AC for the mains by a six-pulse thyristor bridge. The inverter is externally commutated by a synchronous condenser.

The thyristors and diodes are treated as ideal switches. The shaft generator and the synchronous condenser are synchronous machines. Their dynamic behaviour is described in the two-axis transformation. The equations of the other three-phase components are also given.

The stray capacitors, the stray inductance and bleeder resistors can be ignored because they have no
influence on the dynamic behaviour of the plant. They cause an incredible increase of calculation time.

Normally, the coupling of inductive circuits causes trouble for the simulation program and calculation time because the voltages between the circuits are not known explicitly. But in this report a principle is introduced to calculate the voltages explicitly by the currents and the parameters in a simple way. This means a reduction of calculation time so that personal computers can be used to simulate a whole shaft generator plant in an acceptable time.

Bibliography

Fig. 1 Machines of a classic merchant ship

Fig. 2 Ship with a shaft generator plant

Fig. 3 Shaft generator plant with a frequency converter

Main machine

Shaft generator

Propeller

450V/60Hz or 400V/50Hz

Auxiliary machines

3 phase generators mains

MM main machine

SG shaft generator

R rectifier

IV inverter

PS phase shifter, synchronous condenser

CH choke

F harmonic filter

Sw power switch

DB distributing bar

SM starting motor (pony motor)
Fig. 4 Synchronous machine with damper winding 
in the Z-axis transformation

\[ e_d = -\omega \cdot \psi_q \text{ induced voltage } d\text{-axis} \]
\[ e_q = \omega \cdot \psi_d \text{ induced voltage } q\text{-axis} \]

Fig. 5 Three-phase choke in the two-axis transformation
Fig. 6 Three-phase capacitor in the two-axis transformation

Fig. 7 Non-commutation of a six-pulse bridge
T1 and T5 are conducting

<table>
<thead>
<tr>
<th>cycle</th>
<th>u_{12}</th>
<th>u_{23}</th>
<th>u_{31}</th>
<th>l_1</th>
<th>l_2</th>
<th>l_3</th>
<th>conducting thyristors</th>
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<td>1N</td>
<td>u_g</td>
<td>u_k</td>
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<td>T1, T5</td>
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Table 2 Non-commutation cycles of a six-pulse bridge
Fig. 8 Commutation of a six-pulse bridge
T1, T5 and T6 are conducting

Table 3 Commutation cycle of a six-pulse bridge

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<tr>
<th>cycle</th>
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<td>$-i_g$</td>
<td>$i_{gk}$</td>
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</table>

a) conduction through without commutation

b) conduction through with single commutation

c) conduction through with double commutation

Fig. 9 The three types of conduction through
Fig. 10: Synchronous machine with a three-phase bridge and a DC circuit

Fig. 11 Equivalent circuit for Fig. 10

Fig. 12 The DC intermediate circuit between the rectifier and the inverter

SG shaft generator
R rectifier
IN inverter
ES equivalent synchronous machine

Fig. 13 The equivalent DC circuit of Fig. 12 with temporal parameters
Regular falsi:

\[ D_{\text{switch}} = D_t \cdot \frac{i}{i - i_0} \]

\( T_{n, i_0} \) starting values of last integration step

\( T, i \) values of last integration step

\( D_t \) time step of integration

Fig. 14 Calculation of the switching time

Fig. 15 Flow chart of the simulation program
Programm: SMER
Datum: 23.03.78

Simulation eines Synchrongenerators mit 6G-Thyristorbrücke

Daten des Generators:
Xdh= 3.5 Xdp= .15 Xdp= .04 Rdp= .02 Rdp= .02
Xdp= 3 Xdp= .15 Xdp= .08 Rdp= .02 Rdp= .02
Xdp= .05 RF= .885

Daten des Gleichstromkreises:
U1= 0 Rg= 1.7 Xg= .5
Zwundwinkel= 30 Grad

Generator in Leerlauf vor Eintritt der Belastung.
Generator mit Spannungregler, PI-Regler
f= 50 Hz f= 50 Hz const

Plot 1 Synchronous generator with a six-pulse bridge.
three-phase curves

Plot 2 Three-phase voltages and the voltages of
the two-axis transformation
Program UNESC

Datum: 22.03.98

Simulation eines Synchronegenerators mit 98-Thyristorbrücke

Daten des Generators:
Xqe = 3.5  Xse = 1.5  Xq = 0.84  Rq = 0.82  Rq = 0.82
Xiae = 3  Xsa = 2.15  Xae = 0.99  Rse = 0.8  Rq = 0.82

Daten des Gleichstromkreises:
Ug = 0  Rg = 1.7  Xpg = 5

Zündwinkel = 30 Grad

Generator im Leerlauf vor Eintritt der Belastung.
Generator mit Spannungsgeregler, P1-Regler

f = 50 Hz  f' = 50 Hz const

Plot 3: Current of line 1 and the transformed stator currents and the currents of the damper windings

Program US-MS

Datum: 25.07.15

Simulation einer MS-Wellenlastanlage mit Thyristorwechselrichter, Phasenschiebermaschine und Netzlast sowie Kurzschluß des Netzes

Phasenschieber:
Xiae = 2.66  Xse = 0.80  Xa = 0.92  Rse = 0.811  Rq = 0.811
Xae = 1.44  Xpe = 2.06  Xn = 0.857  Rpe = 0.811  Rn = 0.811
Xna = 1.20  Rn = 0.811
Fp = 60 Hz  F = 50 Hz

Wellenlastgenerator:
Xse = 2.38  Xp = 0.329  Rse = 0.8461
Xae = 0.841  Xa = 0.329  Rse = 0.8461
Xn = 0.320  Rn = 0.8276
F = 14 Hz  F' = 14 Hz

Gleichstromkreise:
Rg = 0  Xpg = 0

Drossel zum Netz:
Xn = 0.13  Rn = 0.82

Netz:
Bis Schaltzeitpunkt I sec:
Xn = 0  Rn = 0
Nach Schaltzeitpunkt I sec:
Xn = 0  Rn = 0

Zündwinkel des Wechselrichters = 180 Grad
Spannen der Zündimpulse bei 1 sec

Plot 4: Shaft generator plant with slow-turning shaft generator
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