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INVESTIGATION OF THE POSSIBILITY TO BUILD A 400KA PULSE CURRENT GENERATOR TO DRIVE A MAGNETIC HORN.

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Investigation of the possibility to build a 400kA pulse current generator to drive a magnetic horn.

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Required performance

The magnetic horn is a coaxial cylinder with an inductance of about 420nH. It has to be pulsed with a current of 400kA-peak amplitude. The current pulse is a sine half wave with duration of 10µs. For this pulse width the resistance is 370µOhms taking the skin effect into account. The pulse width of 10µs is long enough to focus the particle beam that has a duration of about 3µs. During the passage of the beam the current will vary from 350kA to 400kA. This current variation can be tolerated. In order to keep the mechanical and thermal stress of the magnetic horn as low as possible the pulse shouldn't be any longer than the minimum required for operation which is about 10µs. Stress minimisation is of great importance to get an acceptable lifetime of the horn because the pulse repetition frequency is 75Hz. The horn will become highly radioactive during operation. Therefore it will be placed behind shielding walls. The current generator must be at a safe distance and will be connected to the horn by about 10m of transmission line. During operation it might be necessary that the direction of current be reversed. This won't happen often so automatic polarity change is not required. The generator should, however, be designed to allow for an easy change of current direction manually. For example by turning all the switches around in the pulse power circuit and the charging circuit.

Principle of the pulse generator circuit

Simplified description of circuit

The pulse current circuit consists basically of three elements: a capacitor, a switch and an inductive load. Figure 1 When the capacitor is charged and the switch is closed a sinusoidal current flows.

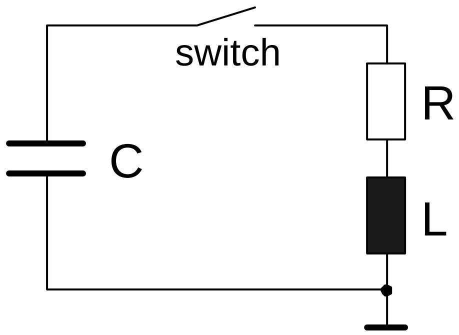

Figure 1

An approximate calculation gives the following results for the capacitor charging voltage and capacitance. The energy stored in the capacitor is equal to the energy in the magnetic field of the circuit inductance.

$$\frac{1}{2}C \cdot v^2 = \frac{1}{2}L \cdot i^2$$

The frequency of oscillation is determined by the following equation:

$$f = \frac{1}{2\pi\sqrt{LC}}$$

With L given as 420nH, the peak current i as 400kA and the period of the oscillation as 20µs the resulting capacitor charging voltage is 52.8kV and the capacitance is 24.1µF.
In this calculation, however, some important quantities have been neglected:

- There is some resistance in the circuit. This will dampen the oscillation.
- The circuit inductance is not concentrated in one component but distributed over the whole circuit. In addition to the inductance of the magnetic horn there is the inductance of the transmission line, the switch and the capacitor.

Though these quantities are difficult to determine it can be assumed that in a practical circuit they are of great importance. The result of this is that in reality the capacitor charging voltage will have to be considerably higher than the voltage of the ideal circuit and the capacitance will be smaller. The stored energy of the circuit will be greater as well because it increases proportionally with the inductance.

How the circuit works

In the ideal case the circuit would operate as follows: At t=0 the capacitor is charged to V_0 and the switch is closed. A sinusoidal current starts to flow. Since only the first half wave is needed the switch opens at the first zero crossing of the current at t=10µs. At this point the capacitor is reverse charged to $-V_0$. The switch has to block a voltage with the opposite polarity as before. Now a recharging circuit charges the capacitor up to the initial voltage and the circuit is ready for the next pulse. The recharging time can be several milliseconds because at 75Hz pulse repetition frequency the period is 13.33ms.

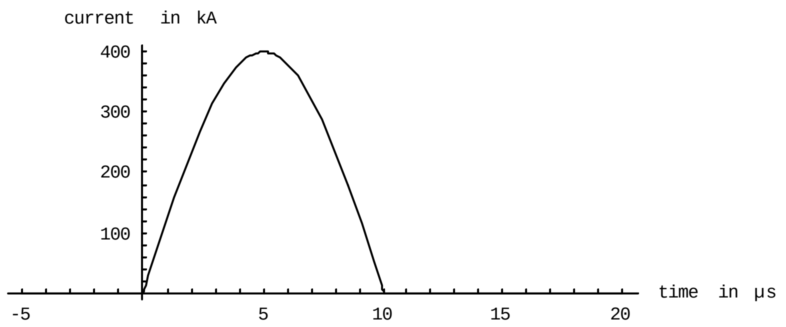

Figure 2a

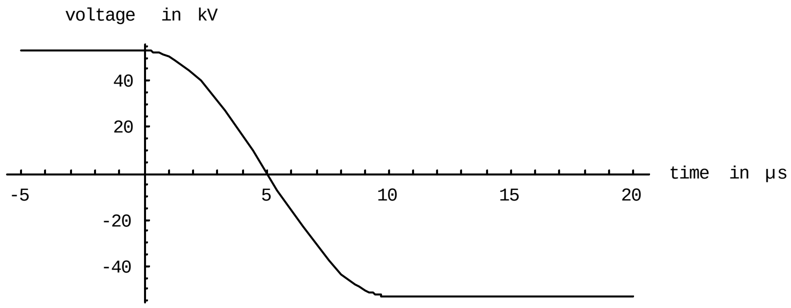

Figure 2b

Figures 2a and 2b show current and capacitor voltage during the pulse in the ideal case.

The switch technology

The expected performance of the switch

Whether such a circuit can be realised or not depends entirely on the available switch technology. The points that are critical for the switch can be summed up as follows:

- High turn on speed. Full blocking to full conductivity in less than a microsecond.
- The switch must tolerate high current rise speed. In this case di/dt = 126kA/µs
- High peak current capability. Because the current is pulsed the peak current is much higher than the RMS current.
- High forward blocking voltage. If several switches with lower blocking voltages are switched in series the circuit inductance is increased and an even higher voltage will be needed.
- High reverse blocking voltage. As high as forward voltage.
- High turn off speed in zero crossing of current. A delay would cause reverse current to flow through the circuit.
- Fast rise of reverse voltage. After the zero crossing the reverse voltage is instantly applied to the switch.
- Assuming that the switch has some losses operation at 75Hz must be possible without causing thermal overload.
- Long lifetime. At 75 Hz the number of pulses in one year of continuous operation are $2.365*10^9$. This means that gas discharge switches can’t be used because their lifetime is far too short.

What is not critical:

- Forward voltage rise. The capacitor charging time is several milliseconds. Therefore the forward voltage across the switch rises slowly compared to currents and voltages during the pulse.
- The forward blocking recovery time. After the pulse current has stopped it will take at least half a millisecond before forward voltage is applied to the switch again. This should be enough time for existing switches to recover.
- Conduction losses. The voltage drop across the switch will be small for most switch technologies compared to the total voltage to be switched. A few 100V more or less don’t matter.
- Off state leakage current. As long as the power loss due to leakage doesn’t reach tens of kilowatts it doesn’t matter.

Commercially available switches that could be used

The ideal switch doesn’t exist. The best that can be done is to try to build a switch out of commercially available components. Here is a list of pulse power switches with a description of their performance limitations. The voltage to be switched is assumed to be 100kV. This is a more realistic assumption that takes into account the whole circuit inductance.

- **Spark gaps**. Used a lot in pulsed power applications. They are fast and can switch high voltages and current. Because of their short lifetime use is restricted to applications with low repetition rates of e.g. one shot per minute. Cooling would also be a problem. Expected lifetime at 75Hz would be about half an hour.
- **Ignitrons**. Too slow, lifetime too short like spark gaps.
- **Thyratron**s. They are fast enough and can switch the voltage required. An array of 20 to 40 in parallel would be needed to switch the current. One year of operation would be possible but then gradual replacement of tubes would be necessary. The price would be high, more than 1MCHF. Cooling by oil immersion would be required.
- **IGBT**. High power transistor. Available for MW of switched power. Biggest devices can switch 3.3kV and 1.2kA average current. The pulse current is limited to a maximum of 2.4kA for this device. To switch 100kV and 400kA 40devices would be needed in series and this 200 times in parallel. The total number of devices would be 8000. This solution would be very expensive. One IGBT costs about EUR1000 and 8000 cost EUR8M. The problem is that the ratio of peak to average current is too small. IGBTs can only block voltages in forward direction. They switch on and off in about 100 to200ns. That would be fast enough.
- **Dynistors**. A Russian invention. Available at the Joffe Institute at St. Petersburg. The person to contact there is Prof. Igor Grekhov. Dynistors are fast high power switches that are similar to thyristors. They are especially designed for pulse power applications and promise to offer good performance. According to the information provided by Prof. Grekhov the biggest devices can switch more than 200kA-peak current. They tolerate a di/dt of up to 60kA/µs and block a forward voltage of up to 3kV. They can’t block reverse voltages. To switch 100kV and 400kA at least 3 parallel stacks of 40 devices in series would be needed. The price per dynistor is about $600. 120 devices would cost $72000. A stack of 40 dynistors with water cooling heatsinks between them would be more than 2m long. This would add a considerable amount of inductance to the circuit. To reduce inductance more than three stacks might be needed in parallel. Dynistors are bipolar devices. They haven’t got a gate connection. To turn them on they have to be revere biased for a microsecond and a reverse current pulse of several 100A has to be injected. See Figure 3. After that they are fully turned on and can immediately conduct large forward current. To allow the short voltage reversal a saturable inductor has to be put in series with the dynistor. Prof. Grekhov has not yet built a triggering circuit for 100kV but he thinks it would be possible. What this circuit would look like is so far a Russian secret. Dynistors are also called RSD (Reverse Switching Dynistor) because of the way they are triggered.

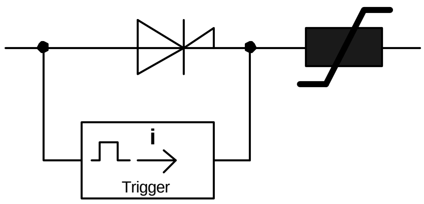

Figure 3 shows a complete dynistor switch with saturable inductor and trigger circuit

- **Thyristors**. Conventional thyristors are far too slow for this application. With critical di/dt < 1kA/µs several hundred would be needed in parallel. Turn-on times of several microseconds are too much.
  Fortunately some manufacturers have developed special high speed pulse power thyristors in recent years. These thyristors are optimised for fast turn on and high di/dt. They can’t block reverse voltage. Because of their special gate structure they require high gate trigger currents of typically 100 to150A for a few microseconds. Examples of these thyristors are:

Manufacturer	Type	Forward Voltage	Peak current	Critical di/dt	On state resistance
DYNEX	PT85QWx45	4500V	90kA	22kA/µs	0.3mΩ
ABB	HCT3001-25A01	2500V	150kA	20kA/µs	-

This list is not exhaustive but the numbers show that it is possible to build a 100kV/400kA switch with these thyristors. Because of the special nature of this project a close collaboration with the manufacturer of the components would be required in the design of the switch. With the Dynex thyristor 6 stacks in parallel with 30 elements in series would be sufficient. Assuming a price of CHF2000 per thyristor the price for 180 of them would be CHF360000.

Conclusion

For switching the current of 400kA there are two realistic possibilities. The dynistor and the pulse power thyristor. These devices have standard disk cases and can easily be assembled to stacks for higher voltages. Cooling is not a problem. The stacks can be built with water cooling heatsinks between the switch elements. Water cooling with de-mineralised water is a much used and proven cooling technology in power electronics. The voltage ratings of the switches show that dynistors go up to 3kV and thyristors up to 4.5kV. Therefore a thyristor stack can be shorter. This means lower circuit inductance. The dynistors need a saturable inductor in series with them. This also increases the inductance of a dynistor stack. Thyristors don’t need that inductor. Triggering a 100kV thyristor stack is no major technical problem. A 100kV-dynistor stack with triggering system has not yet been demonstrated but it should be possible to build one. Dynistors do have an advantage in switching speed, peak current and di/dt but in this application thyristors would be fast enough as well. The cost of a dynistor switch will probably be lower. Given the information so far there is no clear winner. To get reliable information about performance it would be necessary to build a scaled down version of the pulse circuit and test the different devices in it.

Estimating the switch inductance

The inductance of the switch is a critical parameter in the design of the circuit. It should be as low as possible. Low inductance is achieved by placing the current return conductor(s) as close as possible to the thyristor stack. With a voltage of up to 100kV this is a difficult problem because of the necessary high voltage insulation. What can be achieved depends on the construction of the stack and has to be discussed with the manufacturer. One possibility is to place the stack parallel to a ground plane with a certain distance in between. Figure 4

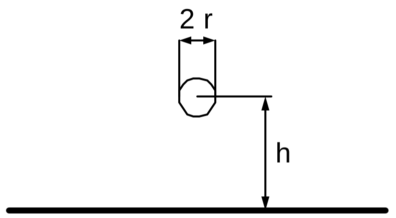

Figure 4

A round conductor with length l radius r and at a distance h over the plane has an outer inductance of:

$$L = \frac{m_0 \cdot l}{2p} \ln\left|\frac{2h - r}{r}\right|$$

The inner inductance is neglected.

For *h* a value of 0.3m is a reasonable approximation. This allows for some insulation material to be put between the plane and the stack and leaves some space for the clamps that hold the stack together. The radius *r* of the conductor is assumed to be the radius of the silicon chip, which is about 40mm. With these values the outer inductance for one meter is 528nH. A thyristor stack with 30 elements and heatsinks would be about 1.5 meters long. Six stacks in parallel would have an inductance of about 132nH.

Another possibility is to design a coaxial structure. In this case four return conductors would be placed around the thyristor stack. Figure 5

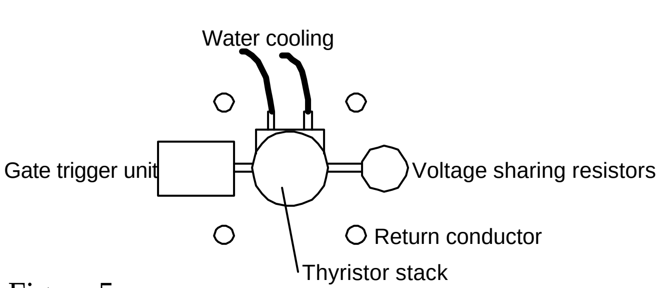

Figure 5

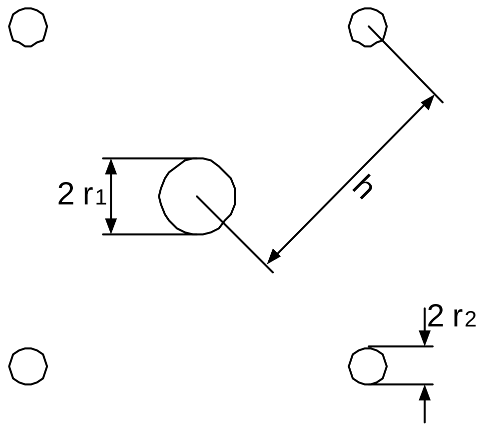

Figure 6

The picture shows a cross section of the thyristor stack. On one side of the stack there is the gate trigger unit on the other side there are voltage sharing resistors. On top of the stack there are the water cooling pipes. In order to get a small inductance the return conductors should be placed close to the thyristor stack. The problem is sufficient high voltage insulation. The inductance is calculated with the following formula: Figure 6

$$L = \frac{\mu_0 l}{2\pi}\left[\ln\left|\frac{h-r_2}{r_1}\right| + \frac{1}{4}\ln\left|\frac{h^4 - r_1^4}{h^4 - (h-r_2)^4}\right|\right]$$

If the current return conductors can be placed close enough to the central conductor an inductance smaller than in case of the current return plane can be achieved. Another advantage of the coaxial structure is that there will be no mutual magnetic coupling between adjacent stacks. In the case of the common return plane there will be some mutual coupling between the stacks that only becomes negligible if the stacks are placed sufficiently wide apart. If possible a coaxial structure should be chosen.

The circuit layout

Option 1

As mentioned in the description of the ideal circuit it would be good if most of the energy stored in the capacitor could be recuperated. During the current pulse the circuit only dissipates a small fraction of the total energy. After the pulse the capacitor is reverse charged to maybe 90% of the initial voltage. The energy can easily be recuperated by the charging circuit. In this case the switch needs forward and reverse voltage blocking. When the current goes to zero after the pulse the switch must open very fast to avoid reverse current to flow. When the switch has opened reverse voltage is instantly applied to the switch. Figure 7 shows how this switch could be built using a thyristor in series with a diode for reverse blocking.

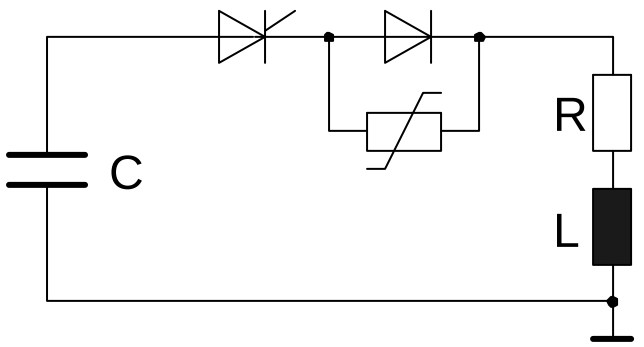

Figure 7

The schematic shows the simplified circuit. L comprises the total circuit inductance. The problem with this switch is that if it were built with real components it would be quite long. Fast diodes are only available with blocking voltages half as high as those of the thyristors. The thyristor stack would be 1.5m long and the diode stack 3m. Six of these stacks in parallel would then have an inductance of (4.5m * 528nH/m)/6 = 396nH, using the result of the previous chapter. Assuming that the transmission line connecting the pulse generator with the magnetic horn has an inductance of 30nH and the capacitor of 100nH the total circuit inductance would be
L_{tot} = 482nH + 420nH + 100nH + 30nH = 946nH.
The result for the capacitor size and voltage would be: C = 10.7µF, V_0 = 119kV. The stored energy would be 75.8kJ. The voltage of 119kV means that the stacks would actually have to be even longer than assumed. Should 100kV be enough then the switch inductance must be 250nH. This would require 10 parallel stacks. Maybe the switch inductance can be reduced by different design but a drastic reduction can not be expected.

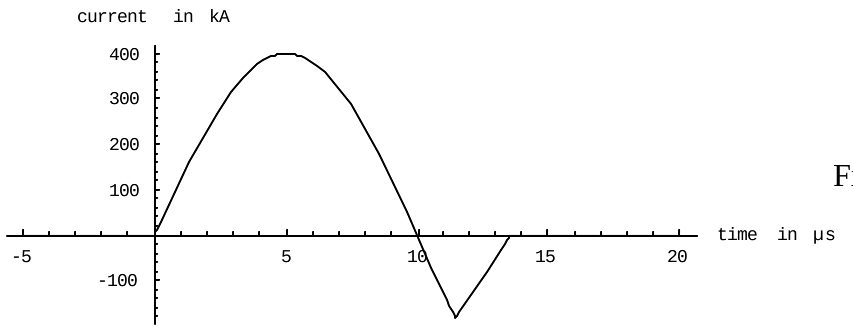

Figure 8a

Current through horn
The RMS current at 75 Hz pulse frequency is 7746A neglecting the reverse current.

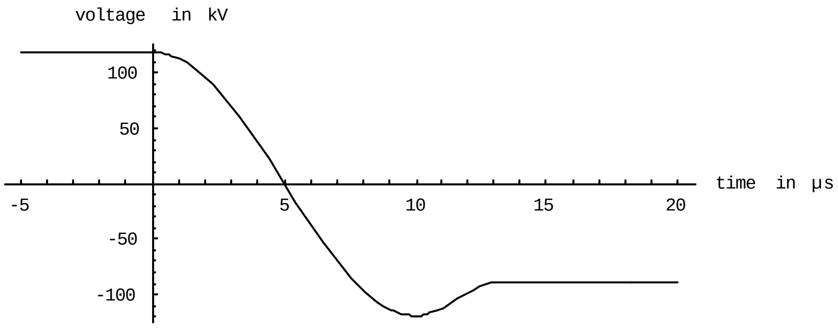

Figure 8b

Voltage of capacitor

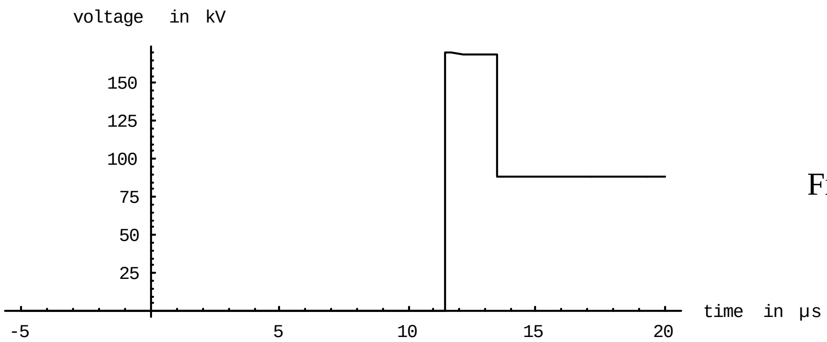

Figure 8c

Reverse voltage across diode

Figures 8 a to c show the current, the capacitor voltage and the reverse voltage of the diode. Because the diodes have a certain reverse recovery time a reverse current will flow for a certain time t_{rr} .In this case t_{rr} assumed to be 1.5µs. The reverse current reaches –180kA. The capacitor reverse voltage is –88kV. With an initial voltage of 119kV the energy recuperated is approximately 55%. That means that in this case about half the stored energy gets lost. The diode has to block a reverse voltage that is much higher than the initial voltage in this case about 50% higher. As a consequence the stack has to be even longer again. A realistic assumption would be a stack with 90 diodes that would be 4.5 m long. Whether this circuit would work or not depends on the available diodes. In order to get reliable data a test circuit should be built and different diodes should be tested in it. The data available from datasheets don't provide reliable information about operation at such high power and speed levels especially about the turn on and turn off behaviour. For this circuit he component cost would be high. This can only be justified if a substantial amount of energy could be saved. If that is not the case a simpler circuit should be built that dumps all the energy. If the circuit is completely de-energised after the pulse the turn off behaviour of the switch components doesn't matter anymore.

Option 2

The simplest way to build a circuit that dissipates all the energy would be a LCR circuit as shown in Figure 9. The resistor R has to be chosen such that the LC oscillator is nearly critically damped.

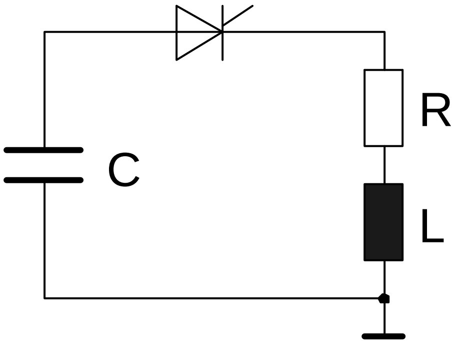

Figure 9

The current in the circuit is described by the following differential equation:

$$\ddot{i} + \frac{R}{L}\dot{i} + \frac{1}{LC}i = 0$$

For critical damping the solution of the equation is:

$$i(t) = \frac{V_0}{L} t e^{-dt} \qquad \text{with} \qquad d = \frac{R}{2L}$$

The condition for critical damping is:

$$R = 2\sqrt{\frac{L}{C}}$$

The current reaches is maximum at $t = 1/d$

This time is assumed to be 5µs so that the rise time is the same as for the undamped circuit. If the switch consists of six stacks of 1.5m length the inductance of the switch is estimated to be 132nH. The total circuit inductance is then 682nH. With a required peak current of 400kA the capacitor charging voltage is V_0 = 148kV. The resistance is R = 0.273Ω. The capacitance is C = 36.6µF. The stored energy of the capacitor would be 400kJ. If this energy were dissipated 75 times per second the total power consumption would be 30MW. This power consumption is probably not acceptable. This pulse circuit has the advantage of a simple design and a small number of components though the 30MW resistor and the capacitor would be quite big. A 30MW capacitor charger would be expensive to build as well. A comparison of the stored energy of 33.6kJ in the magnetic horn with the total stored energy of 400kJ shows that only a fraction of less than 10% of the energy is actually used. This circuit is extremely wasteful. The reason for this is that the strong damping starts a t = 0 and at t = 5µs already 130kJ have been dissipated. The plot of the current Figure 10 shows that the rise time is good but the fall time is very long. This increases the RMS current and is also a reason for the high power consumption. The performance could be slightly improved by not exactly critical damping but allowing small oscillations. The power consumption could be reduced by a few megawatts and the fall time would be a little shorter. But that does not overcome the principle disadvantage of this circuit the enormous waste of energy.

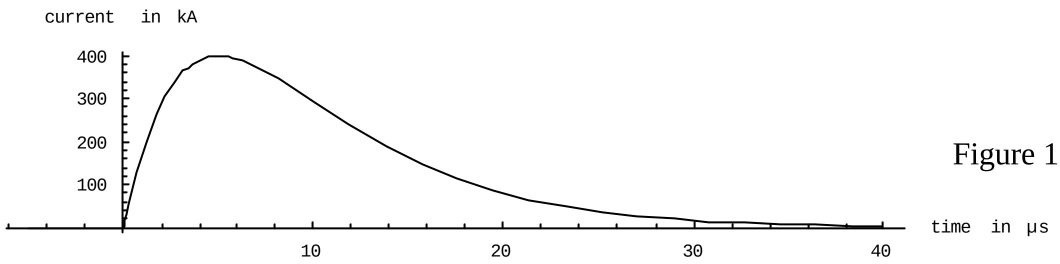

Figure 10

Option 3

If the current could rise without much damping and damping starts after the current has reached its maximum a circuit could be built that dissipates not more energy than necessary to energise the circuit inductance. Figure 11 shows the circuit.

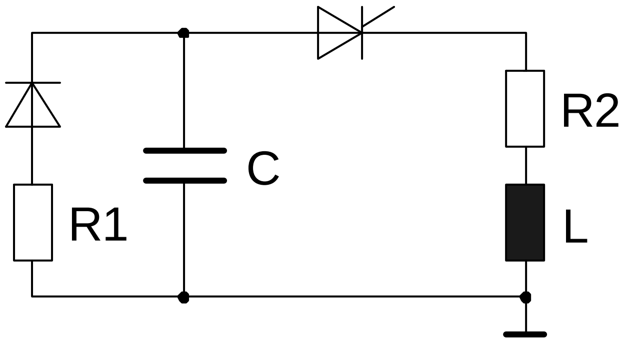

Figure 11

In this circuit the current rises to its peak value with little damping that is caused by the sum of the circuit loss resistances R_2. This is similar to the ideal circuit. After the current has reached its maximum the capacitor voltage turns negative and the diode is forward biased. Now the circuit is critically damped with R_1. The energy is quickly dissipated, faster than in option 2. The diode in this circuit is not very critical. Because of the strong damping the peak forward current of the diode is lower than the current through the horn and the turn off behaviour doesn't matter because the diode is forward biased until the circuit contains no more energy. The plots in Figure 12 a to c show the currents and voltages of the circuit.

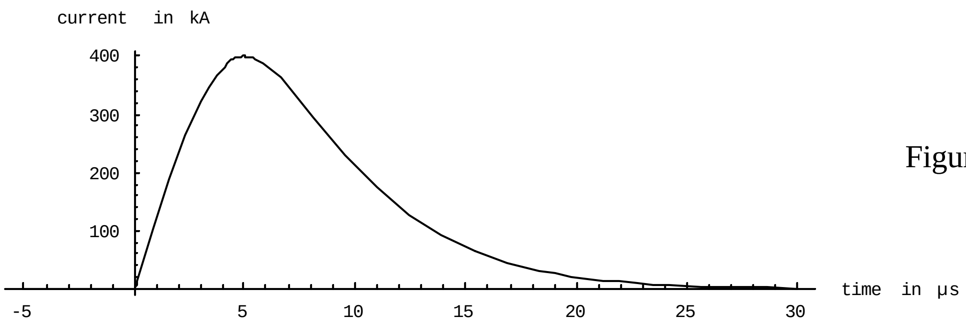

Figure 12a

Current through magnetic horn

The RMS current with 75Hz pulse frequency is 8822A.

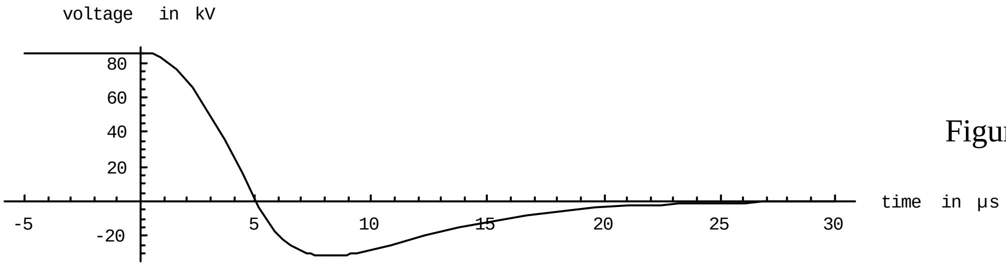

Figure 12b

Voltage of capacitor

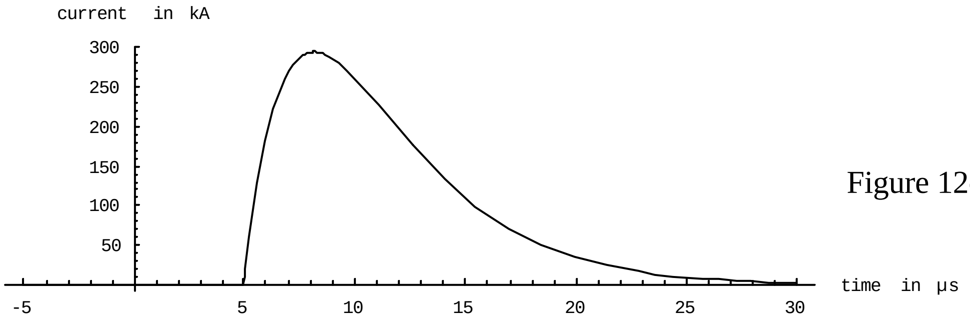

Figure 12c

Current through diode

In this circuit the total inductance is again assumed to be 682nH. To get a peak current of 400kA the capacitor has to be charged to 85.7kV. The size of the capacitor would be 14.9µF. The stored energy is 54.7kJ. If this energy is dumped 75 times per second the total power will be 4.1MW. This is the power that will be dumped in R_1. The current through the horn follows the function:

$$i(t) = i_0 \sin wt \qquad \text{for} \qquad 0 \leq t < 5ms$$

and

$$i(t) = i_0 e^{-d(t-5ms)} \left[1 + d(t - 5ms)\right] \qquad \text{for} \qquad t \geq 5ms$$

With i_0 = 400kA and critical damping for t S 5µs.

$$w = \frac{1}{\sqrt{LC}} = \frac{2p}{20ms} \qquad\qquad d = \frac{1}{2RC}$$

The condition for critical damping with a parallel resistor is:

$$R_1 = \frac{1}{2}\sqrt{\frac{L}{C}}$$

In this case R_1 = 0.107Ω.

This circuit poses the least problems of the three alternatives. The requirements for the components are the lowest. The capacitor is the one with the least stored energy. The voltage is lowest therefore the length of the thyristor and diode stacks is smaller. Component costs can be expected to be lowest for this circuit. The power loss of 4.1MW is still manageable. Given the requirements for switching power and speed this circuit is probably the best solution that is possible with the available switch technology.

Option 4

This circuit presents a different approach that might be of theoretical interest though not of practical one because of the limitations of available switches. So far closing switches have been considered. It is, however, possible to build a circuit using an opening switch. And this circuit could be very energy efficient and would need a lower voltage than the previous ones. It could be built using several thousand IGBTs and would cost several MCHF. The circuit is shown in Figure 13.

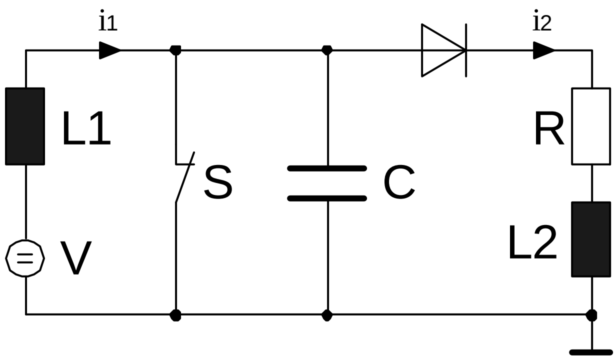

Figure 13

Principle of operation

The power supply V charges the inductor L_1 up to a certain current i_1 about one millisecond before the pulse. This current flows through the switch. This will cause a certain voltage drop across the switch therefore a diode is in series with the load to prevent the current from flowing there. The forward voltage of the diode must be higher than the voltage drop across the switch. The charging current i_1 can be smaller than the required peak load current i_2 .

$$\hat{i}_1 = \frac{L_1 + L_2}{2L_1}\hat{i}_2$$

With $L_1 = L_2$ the currents would be the same. With $L_1 >> L_2$ the current i_1 would approach $i_2/2$. Therefore by making L_1 just a few times greater than L_2 the current could be reduced to about 70%.

When L_1 is fully charged the switch opens and a current i_2 starts to flow.

$$i_2(t) = \frac{\hat{i}_1}{w^2 C L_2}(1 - \cos wt)$$

For the current i_1 the function is:

$$i_1(t) = \hat{i}_1\left[1 - \frac{1}{w^2 L_1 C}(1 - \cos wt)\right] \qquad w = \sqrt{\frac{1}{C}\left(\frac{1}{L_1} + \frac{1}{L_2}\right)}$$

The voltage at the switch is:

$$v(t) = \frac{\hat{i}_1}{wC} \sin wt$$

Based on the assumptions L_1 = 1.5μH, L_2 =450nH (10m cable + horn) and a pulse width T=15μs the plots of Figure 14 a to c are obtained:
$\omega=2\pi/T$, L_1 charging current is 260kA

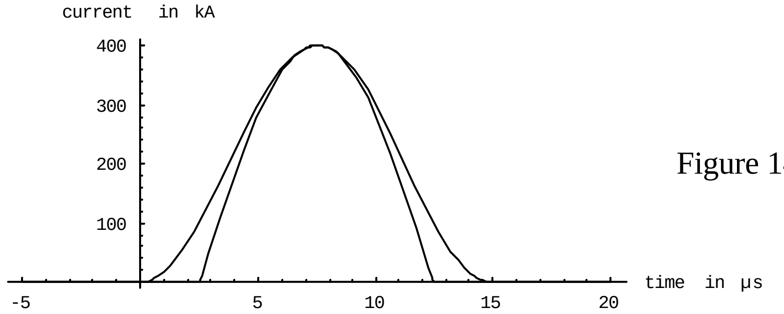

Figure 14a

This is the current i_2 through the magnetic horn. Since the shape differs from the current pulses of the previous circuits a sine pulse of 10μs is shown too for comparison. At the top both pulses are fairly identical.

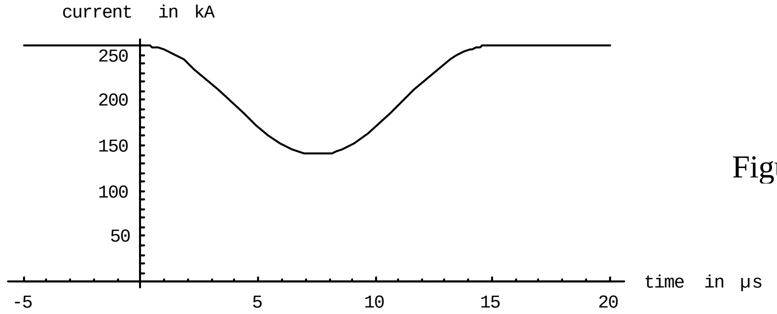

Figure 14b

This is the current i_1 through the inductor L_1 during the pulse.

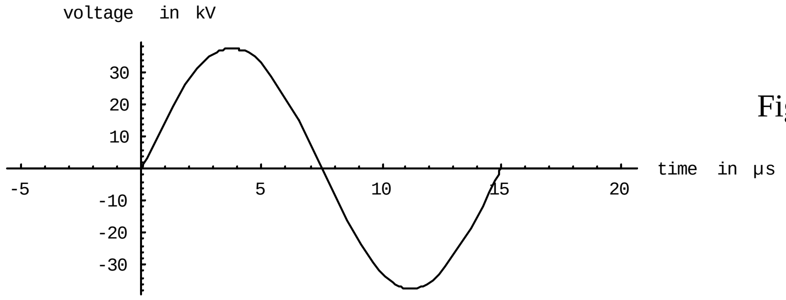

Figure 14c

This is the voltage across the switch. The peak is just at 37.6kV. This is low compared to the other circuits and due to the slower rise of the current.

At the end of the pulse at t=15μs the switch closes, the voltage drops to the low on state voltage of the switch and the L_1 charging current is back to its original level. Now the inductor L_1 can be discharged in less than 1ms by the power supply V. A clever design of the power supply ensures that most of the energy can be recuperated. Another advantage of this circuit is that the peak current of 400kA flows only through the load. The other parts of the circuit see only the current of 260kA.

The capacitor charging circuit

All the pulse circuits of the previous chapter need a capacitor charging circuit. The charger has to charge a capacitor to a certain voltage from either zero or a voltage with opposite polarity. There are several possibilities to build a capacitor charger.

Option 1

The capacitor is charged from a DC power supply via a resistor. Figure 15

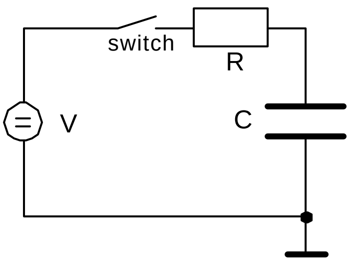

Figure 15

When the switch is closed the capacitor is charged to the voltage of the power supply. At 75Hz operation the time between two pulses is 13.33ms. Leaving 1.33ms for the pulse circuit to recover after the pulse the time to charge the capacitor is 12ms. This is 5 times the time constant RC. The value for R is therefore R = 12ms/(5 C).
The problem with this circuit is that for high charging power the losses in the resistor will be high. In fact when the capacitor is charged from zero to a certain voltage V the energy dissipated in the resistor is equal to the energy stored in the capacitor at V. So in case of the pulse circuit that needs 4.1MW of power the same power would be dissipated in the charging resistor increasing the total power consumption to 8.2MW. This should be avoided. This charger would be no good idea.

Option 2

Figure 16 shows a low loss resonant charger.

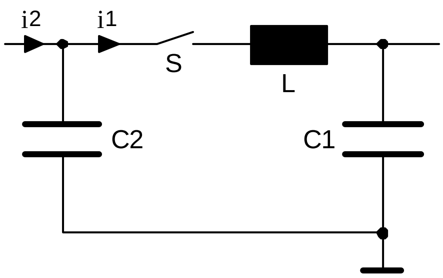

Figure 16

C_1 is the pulse capacitor that has to be charged. C_2 is an intermediate capacitor that stores the energy that C_1 is charged with. When the switch S is closed a current i_1 starts to flow.

$$i_1(t) = \frac{V_{20} - V_{10}}{wL} \sin wt \qquad\qquad w = \sqrt{\frac{1}{L}\left(\frac{1}{C_1} + \frac{1}{C_2}\right)} = \frac{2p}{T}$$

The voltages V_{10} and V_{20} are the voltages of C_1 and C_2 at the time t = 0 when the switch closes. The switch closes for one half period of the sine wave. At the end of the charging current pulse at t = T/2 the capacitors are charged to the following voltages:

$$V_1(T/2) = V_{10} + 2\frac{C_2}{C_1 + C_2}(V_{20} - V_{10})$$

$$V_2(T/2) = V_{20} - 2\frac{C_1}{C_1 + C_2}(V_{20} - V_{10})$$

The current i_2 is a constant DC current that recharges C_2. The components should be chosen such that the charging current duration is between 1 and 2 ms. C_2 should have a bigger capacitance than C_1 to keep the voltage ripple of V_2 small.
The plots in Figure 17 a to d show the charging current and the capacitor voltages.
C_2 = 45µF, C_1 = 15µF, L = 20mH, charging pulse duration T/2 = 1.5ms, V_{10} = 0, V_{20} = 66.7kV.
Pulse capacitor charging voltage V_1 (T/2) = 100kV.

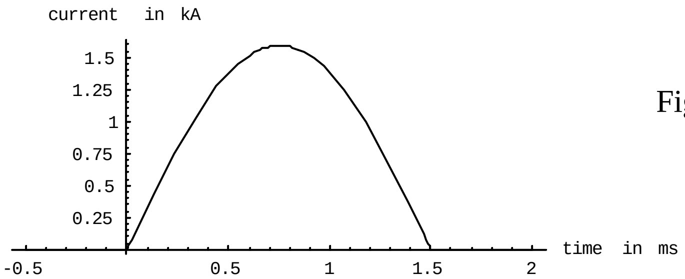

Figure 17a

Charging current i_1(t)

The RMS current of C_2 is approximately 390A for 75 pulses per second. C_2 can be a standard DC filter capacitor. It is subject to far lower stress than C_1.

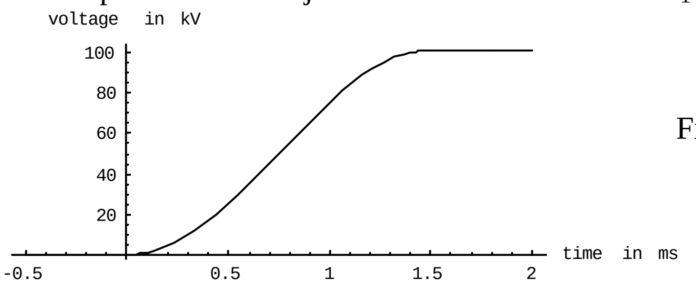

Figure 17b

Pulse capacitor charging voltage V_1 (t)

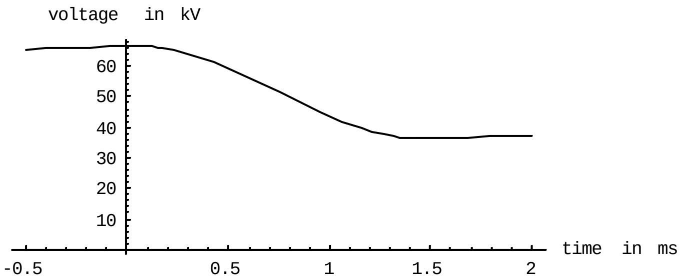

Figure 17c

Voltage of intermediate capacitor V_2 (t)

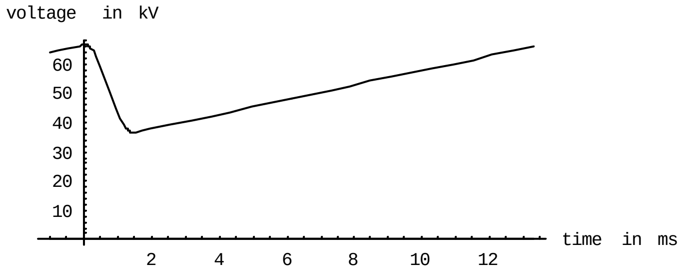

Figure 17d

V_2 (t) over one cycle of 13.333ms of pulse repetition frequency. C_2 is discharged from 66.7kV to 33.3kV in 1.5ms by current i_1(t) and then recharged to 66.7kV by i_2. The current i_2 has to be adjusted by a controlled power supply to recharge C_2 to the correct voltage at the end of the cycle. In this example the current i_2 is about 112A. The average voltage of C_2 is 50kV. The power consumption of the circuit is 112A*50kV = 5.6MW.

The recharge switch *S*

In the example the switch should be a stack of thyristors that are forward and reverse blocking. The requirements for switching speed and peak current are much more modest than those for the pulse power switch. What is critical is the voltage rise dV/dt during the pulse. With 100kV charging voltage the dV/dt is 31kV/µs. With available thyristors that tolerate up to 1kV/µs a stack of at least 31 devices would be needed. The recharge thyristor should be triggered about 1ms after the high current pulse. This 1ms is needed to allow the pulse current switch to recover.
In case of a pulse circuit as described as first possibility that recuperated most of the energy the recharging thyristor stack could be replaced with diodes. The reason for that is that the pulse power switch can recover while the pulse capacitor is still reverse charged. It should take more than 500µs for the voltage to become positive again. A time delay for the recharge switch would not be necessary.

The power supply

The most straightforward way to realise the current controlled power supply is a 6-pulse thyristor controlled rectifier with a big filter inductance L_f. Figure 18. The rectifier output voltage V_{dm} has to be equal to the average voltage of C_2 .

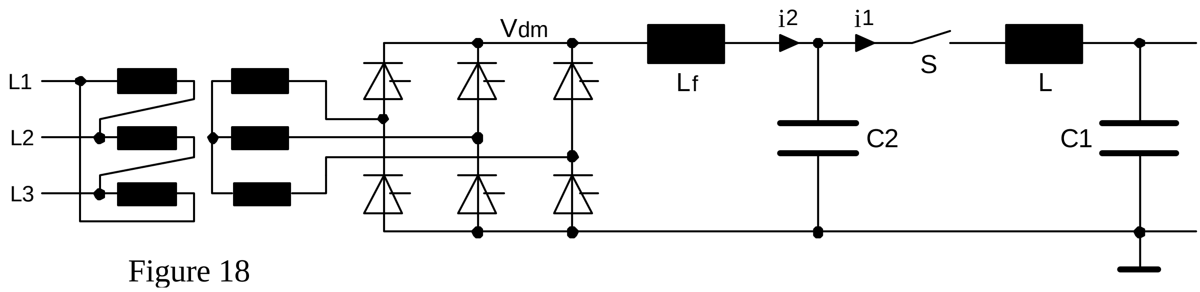

Figure 18

The 6-pulse rectifier generates a 300Hz ripple voltage that has to be filtered by L_f and C_2 . Therefore L_f has to be chosen so big that the ripple voltage of C_2 does not affect the charging voltage of C_1 too much. A big L_f is not a problem for the dynamics of the controller because the load is constant and i_2 never needs to be changed very fast. A further reduction of ripple could be achieved by using a 12-pulse rectifier. This would make the rectifier more complicated but L_f could be smaller. The repeated discharge of C_2 causes a 75Hz ripple current to be added to i_2 . This ripple current also flows in the mains leads. It has therefore to be kept below a certain limit. The ripple is reduced with big enough L_f and C_2 . Another reason for big L_f and C_2 is immunity to fluctuations of the mains voltage. Together with a fast feed-forward control of V_{dm} it should be possible to get sufficient stability of i_2 in case of mains perturbations.

The capacitor charger as described works only under the following conditions:
- The load is constant.
- The pulse current amplitude doesn't need to be changed fast. E.g. different amplitude from pulse to pulse.
- The pulse frequency must not change fast. The circuit can adjust to slow variations over several seconds. Timing jitters and missing pulses must be avoided.

In case of the magnetic horn all conditions are fulfilled.

Conclusion

The biggest problem in building the 400kA pulse generator is the available switch technology. With the dynistor and the pulse power thyristor there are semiconductor switches available that allow construction of such a generator at reasonable cost. High power diodes will be needed as well. It is the performance limitations of these switches that determine what the best circuit layout for this problem would be. In order to get the information needed to make a decision the different devices should be thoroughly tested. Once the characteristics of the switch are known the rest of the circuit can be designed around it.

References

Dynex Semiconductor product guide 2000.

V. Gorbatyuk, I. V. Grekhov and A.V. Nalivkin; "Theory of quasi diode operation of reversely switched dynistors"; Solid-State Electronics Vol. 31, No. 10, page 1483-1491, 1988.

E. Ramezani, A. Welleman and J. Siefken; "High peak current, high di/dt thyristors for closing switch applications"; Ninth IEEE international pulsed power conference Albuquerque New Mexico Vol.2, page 680-683, 1993.

Appendix

Dynistors

The following table shows the characteristics of currently available dynistors.

RSDs Characteristics

	RSD types	TDR1	TDR2	TDR3	TDR4	TDR5	TDR6	TDR7
1	Silicon wafer diameter, mm	76	76	56	40	40	25	25
2	Repetitive Peak Off-state Voltage U_{DRM}, kV	3.0	2.0	2.0	2.0	1.2	2.0	2.0
3	Repetitive Peak Sine current pulse (50 μs width) I_P, kA	250	250	120	60	60	25	25
4	Surge On-state Sine current pulse (10 ms width) I_{ISM}, kA	25	25	12	6	6	2.5	2.5
5	Critical Rise Rate of On-state current (dI/dt) crit, KA/μs	60	60	30	15	15	6.0	6.0
6	Critical Off-state Voltage Rise Rate (dU/dt) crit, kV/μs	0.8	0.8	0.8	0.8	0.8	0.6	0.6
7	Turn-off time tq, μs	250	150	150	100	25	100	15
8	Peak Forward Voltage Drop at $I=I_P$ U_{TM}, V	25	20	20	20	20	20	15

The following pictures show the dynamic behaviour of the TDR3 dynistor. The information can be used to calculate the power loss of the switch. The first plot showing the turn-on process shows that dynistors should be fast enough. For this application the TDR1 dynistor should be chosen. It has the highest blocking voltage, 3kV, and can conduct twice the current because is has twice the chip surface. The price of the TDR2 dynistor is about $600.

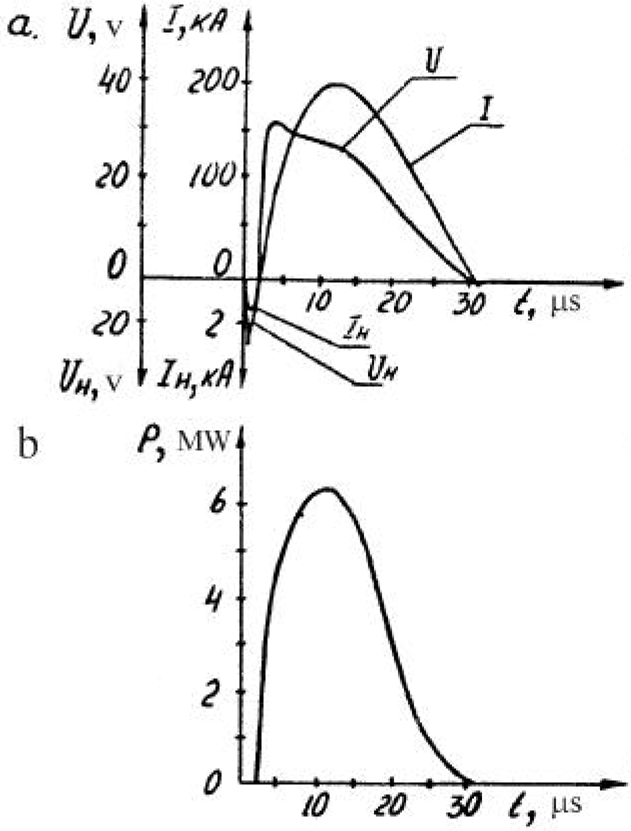

Fig.1. a) Oscillograms of turn-on process in RSD with 20 cm^2 operating surface (56 mm diameter of Si-wafer)
U, I - voltage and current of forward pulse
I_H, U_H - voltage and current of triggering (reverse) pulse
b) Calculated losses

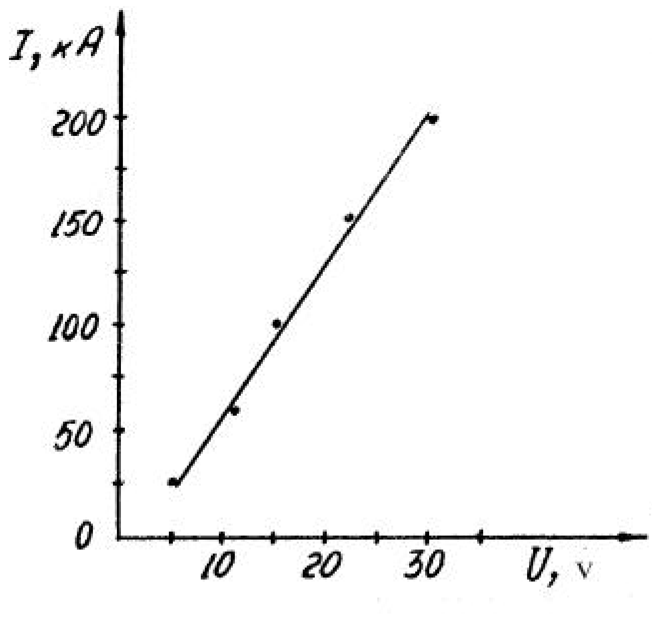

Fig.2. Measured V-I characteristic (at $I=I_{max}$) of RSD with 20 cm^2 operating surface