A microcomputer Controlled Current Source

Danijel Voncina, Rafael Cajhen, Janez Nastran, Ciril Zevnik, Dusan Hudnik
Faculty of Electrical Engineering and Computer Science
Trzaska cesta 25
61001 Ljubljana, SLOVENIA

***Abstract*—This paper describes a prototype microprocessor controlled full-bridge pulse 2.5kW (25V, 100A) Current source that has been designed, assembled and tested. First a brief comparison between dc and pulse plating is presented in the quality of coating. Then the arrangement of each switch, which consists of a parallel connection of eight MOSFET's IRF-150 and its discrete time triggering method is discussed in details. Experimental results, obtained from a prototype current source are presented to illustrate some programming possibilities achieved with the Motorola MC68HC11A8 micro controller and good behavior of the scheme.**

I. Introduction

Pulse plating is a new method for extracting the metal ions from water solutions on the basic material. Modulation of the cathode current with anode pulses is one of possible ways of influencing on several properties of the metal coating. We can say that pulse plating increases the nucleation process, for two main reasons: there is more free energy on the electrode surface than on the DC plating and on the other hand, there exist more possibilities for generating a new nucleus at each new pulse. At this point the use of the pulse current is identical to the adding of several organic additives in the bath.. With the application of this method it is possible to achieve fine grain deposits, low level of porosity, uniform distribution of the coating over the whole surface, improved ductility, etc. [1]. It should be kept in mind that the current "shaping" represents only one way of influencing a complicated system, which contains many other dependent process parameters, such as electrolyte composition, temperature, hydrodynamics and cell geometry. The main purposes of pulse plating are:

- improvement of deposit properties, namely, porosity, ductility, hardness, electrical conductivity;
- improvement of plating thickness distribution by a periodic inversion of the polarity;
- increase in the average deposition rate.

In conditions of DC plating, only average current I can be set. When a pulse current is implemented, several parameters can be set, and there exist numerous rela- tions between each other. The optimum value of each parameter depends on the desired quality of the deposit, as mentioned above. Fig. 1 shows some current shapes that are most interesting in the pulse plating process. The last one is the most complicated and is determined by six time parameters and three amplitude parameters. Since there is no general approach to evaluate parameters, each one should be set in a wide range. The main purpose of this application is to satisfy the demand of researchers in the field of generating pulse current of different parameters in a wide range with the pulse duration from 5 ms to the DC value and amplitudes up to 100 A. It is possible to set nine independent parameters of the pulse current.

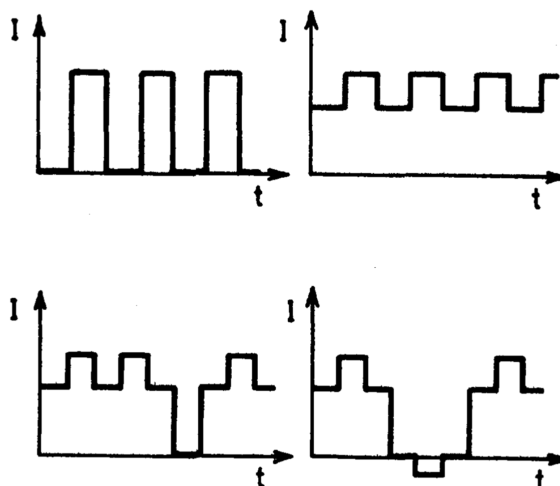

Fig. 1. Different shapes of the pulse plating current

II. Process Current Source

The process current source is designed for an output voltage of 25 V, maximum output current of 100 A, output power of 2.5 kW and maximum switching frequency up to 50 kHz. Fig. 2 is a detailed circuit diagram of the process

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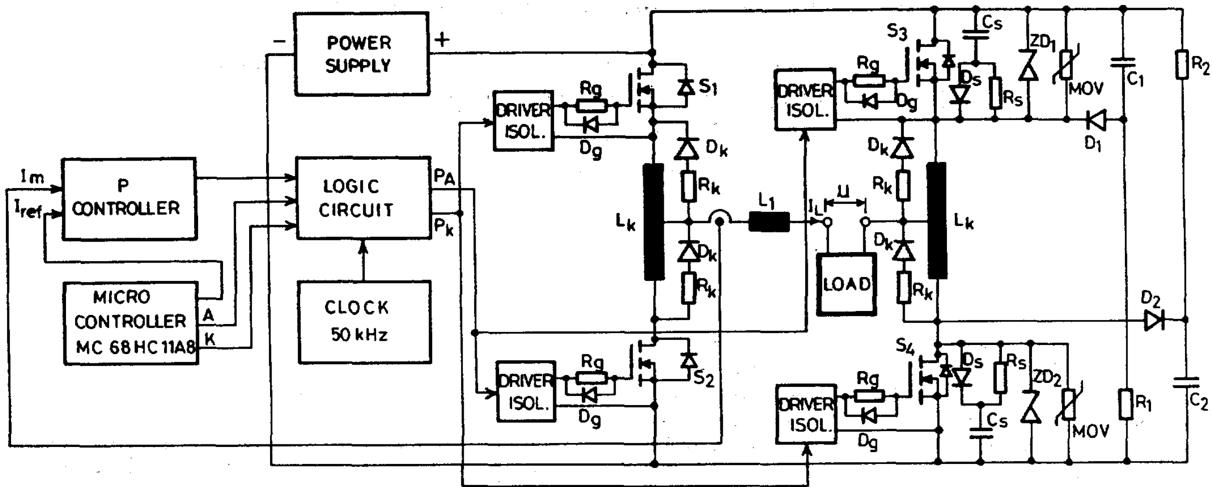

Fig. 2. A Circuit diagram of the process current source

current source, consisting of four switches in the bridge configuration, three phase six pulse power supply, its discrete drive circuits with an isolation stage, feedback and control circuitry and a MC68HC11A8 micro controller, which generates the reference and logic signals for a proper opration of the switches.

III. Circuit Description

A. Switches

The inverter portion of the process current source is made up of four semiconductor switches S_1 - S_4 and their corresponding protection and snubber circuitry.

Each of the four switches consists of a parallel connection of eight MOSFET's IRF-150, which has a voltage blocking capability of 100 V and current rating of 30 A. The uses of MOSFET's have some advantages in this application compared to other semiconductor devices like BJT and IGBT. Being a majority carrier device, there is no inherent delay and storage switching time as that of BJT. The second breakdown effect of the MOSFET is negligible due to the positive temperature coefficient effect. Parallel MOSFET's are used to increase the current rating of the switches. The on-resistance $R_{ds\ (on)}$ of a device is a key parameter that determines the conduction drop. It increases with voltage rating ($\propto V^{2.5}$) making the device very lossy at high currents. The on-resistance of IRF-150 is 0.055 Ω. The resistance has a positive temperature coefficient and therefore permits an easy paralleling of a large number of devices. With eight devices in parallel, the on-resistance and the conduction losses on the switch decrease proportionally to the number of devices. The measured voltage drop on each switch is then only 0.6 V at a 100 A load current. On the same way the inherent parasitic inductance of devices in the switch, which decreases the switching speed, is lower.

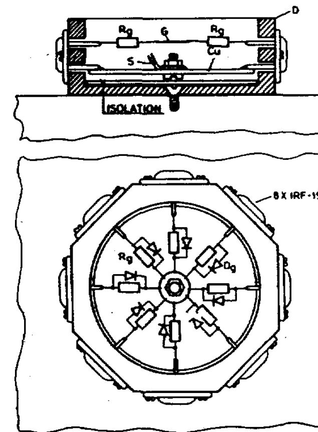

Fig. 3. Configuration of MOSFET's on the heat-sink body

One of the most difficult tasks in designing the power section of the current source is constructing the switch and its integration in the bridge configuration. To ensure fast and simultaneous switching of each MOSFET, some measures must be taken to minimize all loops in the bridge

that draw the currents at the switching frequency. Fig. 3 shows the arrangement of each switch on an additional heat sink body. All interconnections between the MOSFET's and common drain and source connection points are equal and minimal. Such symmetry minimizes stray inductances of drain L_d and source L_s, and represents one step to avoid formation of well-known parasitic oscillations. At the same time there is a low thermal resistance between the additional and main heat sink body, which enables an efficient cooling.

MOSFET's contain an integral antiparallel diode, which may serve as a free wheeling diode, carrying source-to-drain currents when the transistor is off. Although the body diode has a full bypass current capability, it is slow due to a large storage charge. The t_{rr} of the IRF-150's body drain diode is 200 ns. High speed applications require bypassing this diode with external fast recovery diodes [2],[3]. There are two possible solutions to avoid this problem. Usually a series Shottky diode and an ultra fast recovery antiparallel diode are included to prevent the relatively slow body diodes of MOSFET from conducting. In this application, where already two switches draw a relatively large load current in each direction, two additional series diodes with the same current rating as the switch, would not be an economical solution. The second solution contributes less losses in the circuit. The body diodes serve as a free-wheeling diodes. The commutation inductance L_k with a center tap between two series switches slow down the di/dt in the moment when the current for instance commutes between a free-wheeling diode of S_2 and transistor in S_1. The L_k prevents S_1 and S_2, for example, from conducting simultaneously and shortcircuiting the input source. Such situation would be catastrophic to the switches. Free-wheeling diodes D_k enable the release of magnetic energy after each commutation cycle.

The snubber circuits consisting of R_s, C_s and D_s are included to reduce switching stresses during turn-off of the switches. It prevents the switch from high dV_{ds}/dt in case when it acts like an active element and when it works like a free-wheeling diode. This is essential due to a parasitic capacitance between drain and gate, which is V_{ds} dependent. At high dV_{ds}/dt, the C_{dg} may draw a current that leads to raising potential on the gate in case of a relatively high impedance of the drive circuit. The voltage between the gate and source can reach V_{th} and the transistor starts to conduct. High dV_{ds}/dt can result in conducting an internal parasitic BJT of a MOSFET. At high dV_{ds}/dt capacitance C_{db} draws a current and if there is a voltage drop on R_b greater than 0.6 V, the BJT transistor starts to conduct. The RCD circuits reduce dV_{ds}/dt that prevents the transistor from spurious conducting. At the same time the implemented drive circuit, which will be discussed later, contains a protection circuit against uncontrolled conducting due to an internal capacitance C_{dg} of MOSFET.

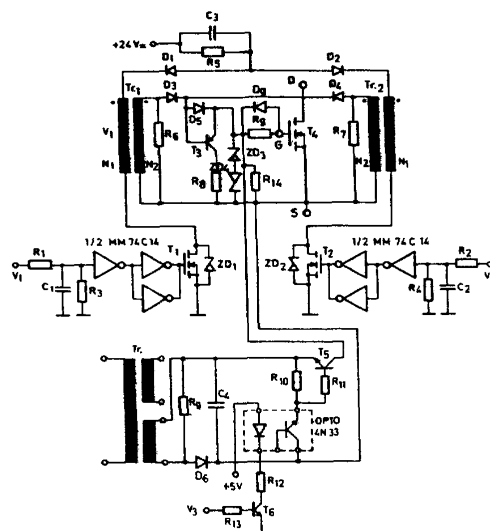

Fig. 4. Drive circuit

B. Drive Circuitry for the Switches

Drive requirements of the MOSFET's are generally less stringent than those of the bipolar transistor. In spite of that, it requires a drive circuit whose complexity may exceed that of the rest of the power section. There are some integrated drive circuits available on the market such as IR-2110 by International Rectifier's and IXBD 4410 / IXBD 4411 developed by IXYS. These circuits have a limited output peak current capability of 2A. They can drive switches with an input capacitances of 10 nF. In this application with eight low voltage MOSFET's, which are known by a high input capacitance, it increases up to 20.4 nF.

This would slow down the switching speed and increase losses. They can not assure short switching times, which leads to increased switching losses. Fig. 4 shows a substitute drive circuit. It includes isolation transformers, protection stage, which serves to protect the switches and driving stage. Transistors T_1 and T_2 turn on alternatively so that pulse transformers Tr_1 and Tr_2 work as amplifiers and provide charge for a fast switch on transition. The switching off transition time depends mostly on the input impedance. PNP transistor T_3, diode D_5 and resistor R_8 provide a low impedance between the gate and source during the switch off transition. The resistor R_g ensures that all eight transistors in the switch turn on and off during the same time interval. As mentioned above, there exists a possibility of uncontrolled turn on of the transistor due to dV_{ds}/dt, which can force the current through the internal capacitance C_{dg}. The voltage drop on the input impedance can reach

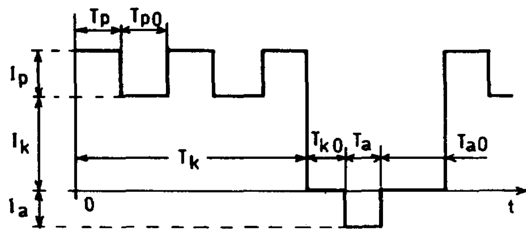

Fig. 5. Parameters of the pulse current

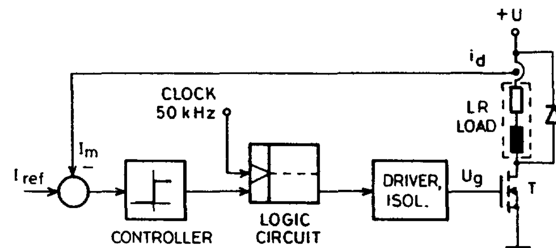

Fig..6. Simplified feedback and control circuit

higher values than U_{th}. This situation can occur when switch S_1 for instance acts like a free-wheeling diode. To avoid this problem, its gate is 3 to 5 V negative biased. The voltage is provided from a separate winding in the drive circuit of the switch S_2 for instance, which is active. This voltage is rectified and filtered by D_6 and C_4. Logic signal V_3 from the micro controller turns on the transistor T_6. The control signal is isolated from the power stage with an optocoupler. Transistor T_5 acts as a switch and enables a negative bias between the gate and source. This measure prevents the situation where switches S_1 and S_2 or S_3 and S_4 turn on and short-circuit the power supply.

C. Feedback and Control Circuitry

The reference signal from the MC68HC11A8 micro controller is fed to the control circuit through a D/A converter. It also generates the control signals to enable switching of S_1 and S_4 or S_2 and S_3. The reference signal is determined by six time parameters and three amplitude parameters as shown in Fig. 5. The micro controller has also some protection functions. If for some reason any of supply voltages is missing, all outputs will shut down, turning off all switches. The output current sensing is achieved using a precision Hall-effect current probe, which provides electrical isolation between the power stage and control circuit. It works on the compensation principle. The information of current is fed back to a control circuitry as a voltage drop on the resistor. Since the current may be either positive or negative, the signal has to be rectified by a precision rectifier that produces a unidirectional signal proportional to the load current. A simplified block diagram of the current regulator is shown in Fig. 6. Measured current Im is compared to reference value I_{ref}. The error is processed only by a P controller that provides a control signal to the logic circuit. The latter synhronizes the control signal with the switching frequency and generates required pulses V_1 and V_2. The power switch can turn on or off every 20 μs. If the load current does not reach the reference value in first the period, it remains open for another 20 μs, and so on, until the reference value is reached. After that, the switch turns off and the load current decreases, until it reaches a lower value compared to a reference signal. After that the switch turns on, and the same procedure repeats.This principle in combination with well-dimensioned load inductance L_1 represents quality current protection of the inverter.

IV. Experimental Results

The performance of the prototype current source was evaluated in the laboratory and in the process of plating. Various waveforms were measured with the Tektronix 2430A digitizing oscilloscope and plotted with a Hewlett Packard 7475A plotter. Some of these waveforms are presented in this section.

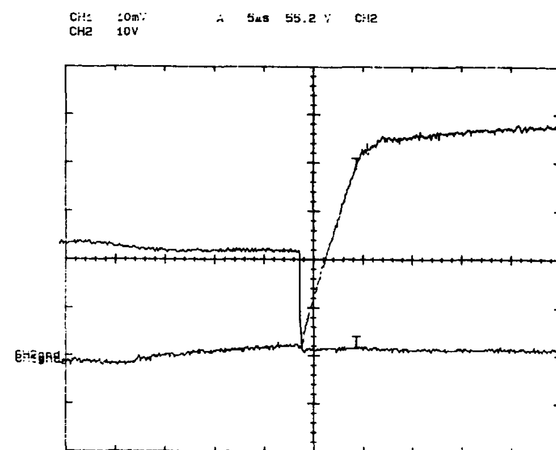

Fig. 7. Turn on of the switch at 100 A load current. U: 10V/div, I:20A/div

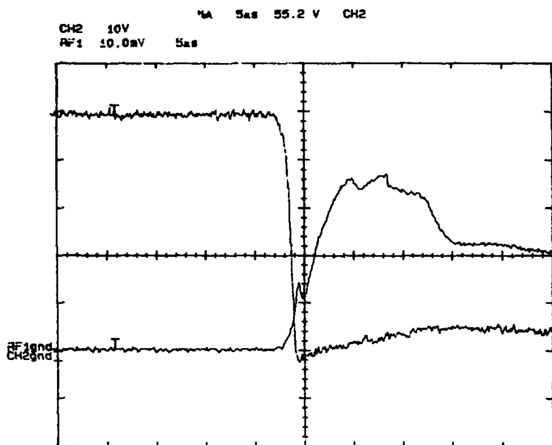

Fig. 8. Turn off of switch at the 100A load current. U: 10V/div, I:20A/div

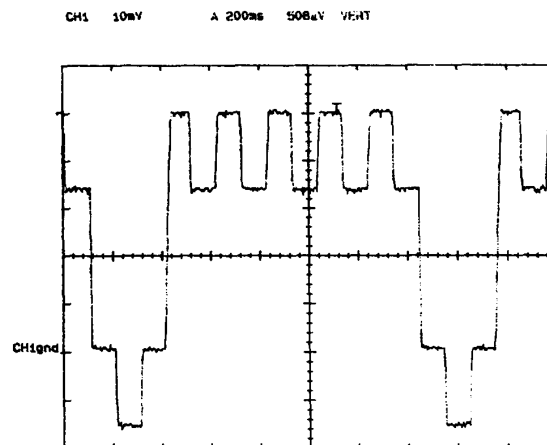

Fig. 9. Load current I:20A/div

V. Conclusions

This paper has described a microprocessor controlled full-bridge current source. The arrangement has been shown to be feasible for use as a pulse current source, which provides current of cathode and anode polarity. The dynamic range of the scheme depends mostly on the supply voltage and the load inductance. A desired load current response can be set by a proper adjustment of these two parameters. Furthermore, the high switching frequency allows low current ripple in the steady state as shown in Fig. 9. A process current source was implemented in the process of pulsed electrodeposition of Nickel. A significant improvement of a deposit structure was achieved, which leads to a reduction of its thickness by 30% and at the same time there were fewer additives in the solution.

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