

## 24 kA solid state switch for plasma discharge experiments

P. Pribyl<sup>a)</sup> and W. Gekelman

*Department of Physics and Astronomy, University of California, Los Angeles, California 90095*

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We present the design of a modular, robust, high-current switch for controlling discharge current in the large plasma device (LAPD) at UCLA. As developed, 10 insulated gate bipolar transistor units are connected in parallel to provide 24 kA current-carrying capacity. Associated snubber and protection circuits are described in detail. The switch is typically operated at a 1 Hz repetition rate, with up to a 2% duty cycle. It has proven to be quite reliable, to date having pulsed approximately  $5 \times 10^7$  times, with a net throughput of about 2 billion Coulombs. This type of switch not only has an application in such a plasma source, but can also be used to switch large currents for a variety of applications. The design is modular and the switch can be made smaller or larger depending on the desired current rating. © 2004 American Institute of Physics. [DOI: 10.1063/1.1646732]

### I. INTRODUCTION

The large plasma device (LAPD) series of plasma devices has been well described elsewhere.<sup>1</sup> Briefly, it consists of a linear magnetized plasma chamber, with a plasma source at one end. The plasma source consists of an 80 cm diam barium oxide coated cathode and a semitransparent wire mesh anode, which are 30 cm apart. A multikiloampere dc discharge is drawn between the anode and the cathode, with the resulting plasma streaming through the anode and down the length of the machine. The length of the plasma column is 18 m, with an axial magnetic field of up to 4 kG. In this article we describe the transistor switch that controls the plasma discharge current. Electrical considerations that form the basis of such a switch are also discussed in some detail.

### II. CIRCUIT OVERVIEW

The circuit used to produce the discharge is shown conceptually in Fig. 1. A capacitor bank stores the electrical energy used to produce the discharge, and this bank is connected between the anode and cathode through a semiconductor switch. The semiconductor switch consists of insulated gate bipolar transistors (IGBTs).<sup>2</sup> Once the switch is energized, current begins to build up in the plasma circuit; a typical wave form is shown in Fig. 2. After an exponential rise phase, this current reaches its desired peak value. Since the voltage across the discharge tends to clamp at between 30 and 40 V, this peak current is determined by lead resistance and capacitor bank voltage. Both rise time and discharge voltage are dependent on working gas and fill pressure. Experimental data are typically acquired in the 2–10 ms “flat top” period during which the current is approximately constant. The IGBT switch is then turned off. The flyback diode serves to prevent the inductive voltage transient that would be caused at this time by the sudden decrease in current flowing in the leads.

<sup>a)</sup>Author to whom correspondence should be addressed; electronic mail: pribyl@ucla.edu

In practice, semiconductor elements in the circuit in Fig. 2 are each composed of a number of parallel modules, since the desired switched current is at least 24 kA, and the largest single-device current capacity for IGBTs at the time of design was 2.4 kA. The current capacity for single diodes was also insufficient. The voltages involved are relatively low, and well within the 1200 V capacity of the transistors.

Energy supplied to the capacitor bank comes from a charging supply connected as shown in Fig. 1. The dc source is a simple three-phase transformer/rectifier circuit, connected to a variac to control the charging voltage; the dc output voltage is settable at up to 150 V, with a current capacity of 100 A. The series charging resistor consists of two 1  $\Omega$  units similar to Powerohm Type ERB4-29, each capable of a rms current of about 30 A, connected in parallel.

### III. DETAILS OF THE SWITCH, CAPACITOR, AND FLYBACK CIRCUIT

The IGBT switch, capacitor bank, and flyback diode, shown conceptually in Fig. 1, are actually composed of ten identical circuits connected in parallel. A number of considerations go into each of these components. The first consideration is the simplicity of driving the IGBTs—the previous LAPD switch consisted of 16 high-current transistors connected in parallel, and required substantial amounts of base current to turn them on as well as special considerations to insure that the transistors turned off rapidly and coincidentally.

One of the main issues in the design of high current switching circuits is the management of voltage transients. If a circuit is conducting 2400 A and turns off in 3  $\mu$ S, even 1.5  $\mu$ H of inductance can result in a voltage transient of 1200 V, enough to destroy the transistor. A loop formed from a 2 m length of No. 1 cable has this much inductance.

Another issue is the sizing of the various cables and components. Since the discharge is pulsed, conductors used in the design need only to be rated for the rms current, and therefore may be substantially smaller than for a steady state system. The target 2% duty cycle allows us to use cables

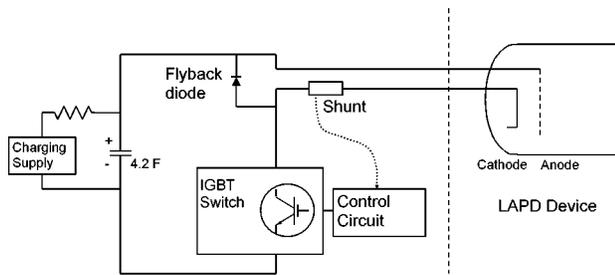


FIG. 1. Conceptual schematic of the capacitor bank, switch, and LAPD cathode/anode circuit. No single IGBT or diode had sufficient current capacity to handle the current, so a number of devices are connected in parallel.

rated to carry  $\sqrt{0.02} \approx 14\%$  of the peak current. This is achieved by using three parallel lengths of No. 1 cable (or two lengths of No. 1/0). For the semiconductors, junction temperature is the limiting factor, and tens of milliseconds is sufficiently close to steady state that not much derating can be gained from thermal heat capacity (although we rely on this 20%–30% safety margin to account for minor variations in wiring and current limits when we use the transistors at their full 2400 A rating).

Finally, operating semiconductors in parallel requires attention be paid to current sharing issues. The saturation voltage of the IGBTs has a negative temperature coefficient, falling with increasing temperature at approximately 2 mV/°C. Thermal runaway can occur if one of the paralleled devices heats up more than the others. This is mitigated in our design by: (1) forced cooling to prevent large temperature differences between devices, and (2) 0.25 mΩ of resistance in series with each IGBT that provides enough voltage drop that differences in collector-to-emitter voltages between devices have reduced effects. Another current sharing problem can arise dynamically: if not all transistors turn off at the same time, the last transistor to turn off will end up carrying all the current. (This issue plagued the previous transistor switch on the former LAPD machine.) We provide for equal propagation delays and use optical isolators having <10 ns time skew along all signal paths to insure identical turn-off times for the IGBTs. Differences in turn-off times are sufficiently small that the IGBT-to-IGBT circuit inductance prevents significant current rise due to this effect.

### A. Individual switch circuits

The switch constructed for the LAPD plasma source at University of California at Los Angeles (UCLA) consists of

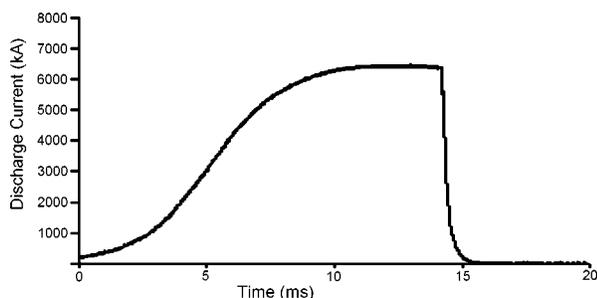


FIG. 2. Typical discharge current wave form in LAPD. The current rise is approximately exponential at first, with a time constant of 1–3 ms.

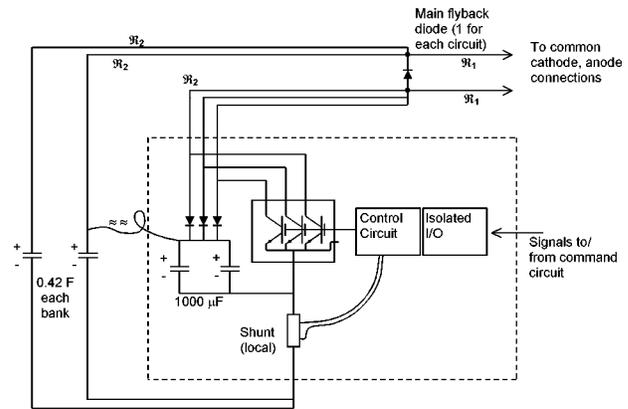


FIG. 3. Combination capacitor bank/IGBT/flyback diode module. Five such modules are connected in parallel within a single cabinet, with two cabinets being used to drive the system for a total current capacity of 24 kA. Components inside the box are all mounted on the IGBT unit itself.

ten identical circuits. Each of the individual circuits consists of two 0.42 F capacitor banks connected to a 2400 A IGBT, together with various snubber circuits, flyback diodes, and control electronics. Figure 3 illustrates one of the ten duplicate circuits that are connected in parallel. At the heart of the figure is the Eupec type FZ2400R12KF4 IGBT. While one of these 2400 A devices costs upwards of \$2000,<sup>3</sup> it is actually cost effective for us to use these instead of a larger number of much cheaper devices due to the reduction in duplicated circuits (e.g., another design at the same current consisted of a total of thirty 400 A devices, for a total semiconductor cost of ~\$7500, but with a much larger amount of associated support circuitry). The 2400 A transistors, with design techniques suitable for insulated-gate devices,<sup>4,5</sup> and with appropriate current limiting have proven to be robust in this application. Since the switch began its operation, it has pulsed approximately  $5 \times 10^7$  times, with a net throughput of about 2 billion Coulombs.

The IGBT unit has three collector connections; we use each of them separately as shown in Fig. 3. Each wire labeled  $\mathcal{R}_2$  is a 2 m length of No. 1 cable, and provides 0.8 mΩ of resistance to help insure current sharing. This redundancy is important in the event that a current monitoring shunt fails (as actually happened during testing): with all current trip points set to 2400 A, circuit-enforced current sharing enables transistor protection using even a single good shunt, since the transistors will all overcurrent approximately together. The current carrying capacity of these wires is sufficient for a maximum duty cycle of 2%, well within the typical operating parameters.

The components inside the dashed box in Fig. 3 are all mounted directly on the IGBT module in order to achieve the smallest possible wiring inductance, as shown in the photograph in Fig. 4. Three snubber circuit diodes (IXYS type MEO500-06DA) are fast recovery units and are connected to a pair of 1000 μF capacitors. The voltage on these capacitors is maintained at a dc value matching that on the capacitor bank by a connection consisting of approximately 1 m of No. 10 wire, designated in the schematic by “≈≈.” At the time the transistor turns off, the local circuit inductance of approximately  $12 \pm 5$  μH in the loop consisting of the mini-

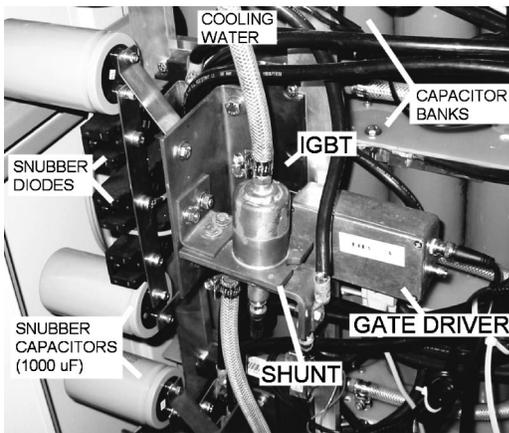


FIG. 4. Photograph of one of the high-power transistors with the associated gate driver, shunt, and snubber circuit mounted directly to it. This photo is of one transistor in a vertical stack of five.

banks and the connections labeled  $\mathcal{R}_2$  tends to keep the current flowing, which if unclamped would cause a damaging voltage transient across the IGBT. The snubber diodes prevent this by becoming forward biased as the collector voltage increases above the supply voltage, allowing the current to transfer to the pair of 1000  $\mu\text{F}$  capacitors. At full current, there is a net transient voltage at the collector of less than 100 V above the supply voltage. This transient includes an inductive component (the voltage around this small loop causes the rapid increase in the diode current) together with a resistive drop inside the capacitors and some amount of rapid charging of the capacitors. This excess charge is then drained off the capacitors by the No. 10 wire over a 25  $\mu\text{s}$  time scale. These diodes conduct only for this 25  $\mu\text{s}$  period at the IGBT turnoff, and may be rated for less than the full dc current. Note that individual connections to the charging supply for each pair of the 0.42 F banks are not shown in Fig. 3, but consist of equal lengths of No. 1 wire connected to the dc supply.

Each of the unit circuits has its own 2500 A flyback diode (type Powerex A451PD). Because of this, the magnetic energy contributing to the above-described transient is only that from wiring on the left side of the flyback diode in Fig. 3. Magnetic energy stored in the machine connections (to the right) does not contribute; rather, at turnoff, the flyback diode also becomes forward biased and limits the contributed voltage drop to approximately 1 V in the  $\mathcal{R}_2$  loop. The flyback diodes are connected to the main leads going to the anode and cathode through wires labeled  $\mathcal{R}_1$ , which are 1 m lengths of No. 1/0 cable. Any magnetic energy stored in these leads and the plasma that is not dissipated in the plasma is thus dissipated in this flyback diode, thereby reducing the load on the snubber.

Each transistor has its own gate drive circuit described in detail below. This circuit is connected to a 0.2 m $\Omega$  coaxial shunt, and will shut off the gate drive if that transistor's current exceeds a present limit. These shunts were constructed in-house, and are described in Sec. III D. The connection to the shunt also provides the emitter reference voltage, so these leads are kept as short as possible.

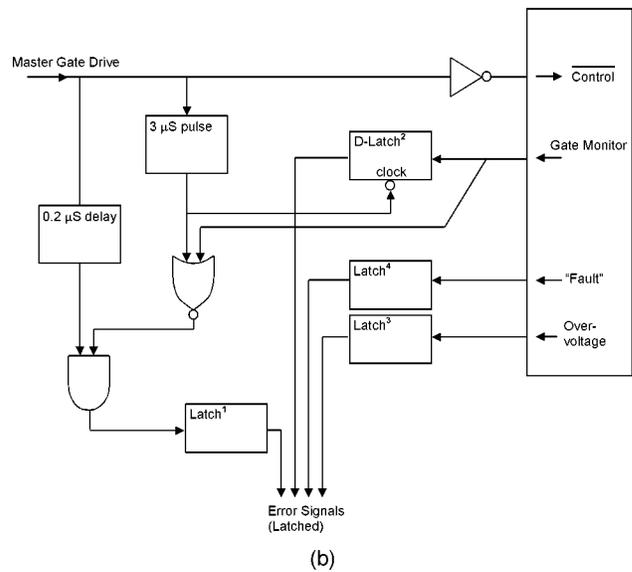
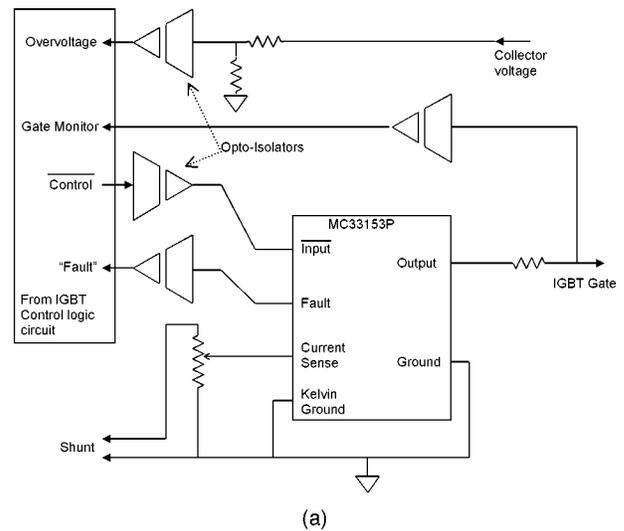


FIG. 5. (a) Gate drive circuit block diagram. This is constructed around the MC33153 circuit, with optically isolated control and status signals. Each circuit has its own isolated power supply. Circuit ground is connected to the emitter of the IGBT through the shunt connection. (b) Block diagram showing the gate control logic for a transistor modules. Each of the circuits in (a) is driven by such a circuit. The 3  $\mu\text{s}$  delay allows time for the gate control signal to bring the gate high, and for the gate monitor signal to propagate back to this electronics. Latch No. 1 detects an overcurrent condition, latch No. 2 detects an absence of a gate drive signal, latch No. 3 detects an overvoltage, and latch No. 4 is the MC33153 "fault" signal (which indicates that even though an overcurrent condition shut off the gate drive the emitter current continued to rise).

**B. Gate drive circuit**

A block diagram of the gate drive circuit for the individual IGBTs is shown in Fig. 5. The gate signal itself is provided by an MC33153 "Single IGBT gate driver."<sup>6</sup> This chip has a local current sense input for monitoring the emitter current of the device being controlled, and internal logic for shutting off the gate drive signal if the current exceeds a particular setpoint. The voltage on the emitter shunt resistor in each of the IGBT circuits (as shown in Fig. 3) is connected across a 100  $\Omega$  trimpot that sets the actual current limit. Since the current sense input trips at 65 mV, the lowest setpoint we can specify is approximately 325 A in the 0.2

m $\Omega$  shunt. If the current continues to increase above this trip point, a second internal setpoint at 130 mV will raise a high signal on the “fault output” pin of the MC33153, indicating a serious problem has occurred in the high-current portion of the circuit. There is no output indicating a simple overcurrent trip, however, so we eventually detect this in the master control circuit by monitoring all of the gate drive outputs and comparing them to the master *gate-drive-on* command. If any transistor’s gate drive shuts off while the *on*-command is active, we register a fault and shut off the gate drive for all transistors.

A pullup resistor, not shown in Fig. 5(a), is connected to the shunt input. If the shunt connection opens for whatever reason—e.g., the shunt center conductor opens, or the local connection to the shunt fails—the “current sense” input will be presented with a permanent “overcurrent” indication, preventing operation of the system.

The gate drive command signal and all output status signals are brought to each individual gate drive circuit using an RS-232 style cable connected to an ungrounded DB9 jack. Once in the gate driver box, these signals are optically isolated via HCPL-7720 integrated circuits. These isolators have a propagation delay of 40 ns, and an internal electrostatic shield between the optical detector and the light-emitting diode rated to provide common mode transient immunity of at least 10 kV/ $\mu$ S. The 5 V dc supply for the isolated stage is provided on the same cable that carries the signals. Power for the gate driver box is provided by a floating regulator circuit connected to a split-bobbin transformer.

### C. 0.42 F minibanks

Each of the 0.42 F banks shown in Fig. 3 consists of 35 parallel-connected capacitors; each capacitor is 12 000  $\mu$ F at 200 V (Nippon Chemi-con type 36DY). The cans are mounted to a sheet of 0.25 in. thick G-10 fiberglass, with each of the negative terminals hard wired to a copper strap situated between the capacitor terminal and the fiberglass. Each of the positive terminals are connected via the mounting screw through the fiberglass to a lug. Copper wires, 0.015 in. in diameter, then connect each of these lugs to a single common positive buss running the length of the fiberglass. In the event that one of the cans should fail, these wires act as fuses limiting the amount of energy that can be transferred from the rest of the bank to the bad one, and thus preventing exploding cans. This technique is based on a similar arrangement used at the UCLA Tokamak Lab<sup>7</sup> and elsewhere. In particular, the  $I^2t$  for one wire is computed to be approximately 2800 A<sup>2</sup>s, which is incidentally independent of the length. At a charging voltage of 60 V, the short circuit current is about 4 kA and will vaporize the wire in under 0.2 ms. Thus, no more than 4 kA $\times$ 60 V $\times$ 0.2 ms $\approx$ 50 J can be transferred to the can in this time, which is not enough to destroy it. Each wire has  $\sim$ 10 m $\Omega$  resistance, so 35 in parallel adds approximately 0.3 m $\Omega$  of series resistance to each minibank.

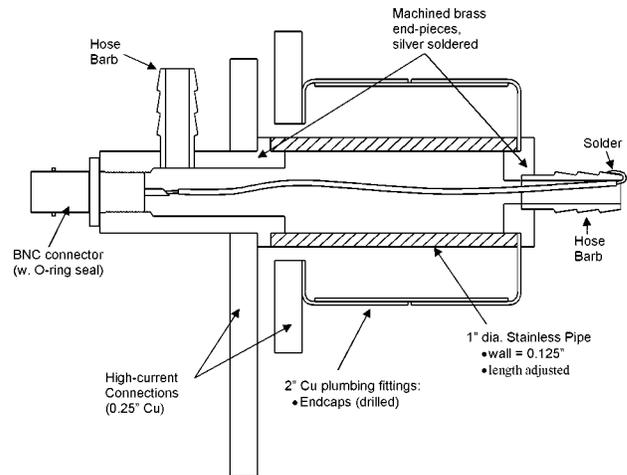


FIG. 6. The coaxial water-cooled shunt. The resistive element is a section of stainless steel pipe, having a diameter of 1 in. and 0.125 in. wall. Its length is adjusted to give a resistance of 0.2 m $\Omega$ , about 2.25 in. in this case.

### D. Coaxial shunts

Each shunt is constructed from a 2 in. length of stainless steel pipe, together with some 2 in. copper pipe fittings and several custom brass pieces. The cross section is shown in Fig. 6. All pieces are silver soldered together, with the exception of the plumbing fittings and center conductor which are soft soldered. The stainless pipe forms the resistive element, and its length is adjusted to give a resistance of 0.2 m $\Omega$ . For the 1 in. diam pipe having 0.125 in. walls that we used, the appropriate length was measured to be approximately 2.1 in. In constructing the shunt, we cut the pipe to 2.25 in. to account for the soldered endcaps. Variability from shunt to shunt is about 5%, primarily due to vagaries of the silver solder process.

Since the temperature coefficient of the resistivity of the stainless element is larger than  $10^{-3}/^{\circ}\text{C}$ , and the average power dissipation may be as much as 100 W, water flow through the stainless pipe is used to maintain a relatively constant temperature. The center coaxial conductor is a piece of No. 10 teflon-insulated wire, with silver-plated strands. We have found this kind of wire to be more resistant to corrosion than others. The time constant for this shunt to measure a step change in current is about 3  $\mu$ s, which corresponds to the current’s diffusion time through the wall of the stainless pipe.

A better design for these shunts should probably have the water flowing on the outside of the stainless, with an insulating seal between the conductors for the two ends. This will mitigate the corrosion observed on the center conductor due to dirty water.

### ACKNOWLEDGMENTS

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<sup>1</sup>W. Gekelman, H. Pfister, Z. Lucky, J. Bamber, D. Leneman, and J. Maggs, Rev. Sci. Instrum. **62**, 2875 (1991) (The LAPD web site is at <http://plasma.physics.ucla.edu/lapd/>)

<sup>2</sup>IGBTs are Eupec type FZ 2400 R12 KF4 (see <http://www.eupec.com>)

<sup>3</sup>Amps Abundant, Division of Silicon Power Corp., 11125 Knott Ave., Suite D, Cypress, CA 90630.

<sup>4</sup>International Rectifier Application Note AN-955, Protecting IGBTs and

MOSFETs from ESD (<http://www.irf.com/technical-info/an955/an-955.htm>)

<sup>5</sup>International Rectifier Application Note AN-949, Current Ratings of Power Semiconductors (<http://www.irf.com/technical-info/an949/an-949.htm>)

<sup>6</sup>ON Semiconductor (<http://www.onsemi.com/pub/Collateral/MC33153-D.PDF>)

<sup>7</sup>R. J. Taylor, UCLA Tokamak Lab (personal communication).