

**ADVANCED  
POWER  
TECHNOLOGY®**

**APT9601**

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# APPLICATION NOTE

## **HIGH FREQUENCY RESONANT HALF BRIDGE**

**MOS-Gated Power Semiconductors Configured in the  
ZVT Thyristor-Dual Mode Yield > 95% Converter  
Efficiency at 1-10 kW, When Resonantly Switched  
at 20-400 kHz**

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# **MOS-GATED POWER SEMICONDUCTORS CONFIGURED IN THE ZVT THYRISTOR-DUAL MODE YIELD > 95% CONVERTER EFFICIENCY AT 1-10 kW, WHEN RESONANTLY SWITCHED AT 20-400 KHZ**

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## **INTRODUCTION**

MOS-GATED SWITCHES, IGBTs AND POWER MOSFETs, by virtue of their ease of control and good dynamic behavior - notably the absence of storage time at turn-off, have opened new perspectives for advanced converter topologies. In particular, detailed analysis of commutation mechanisms over the last few years has led to a greater recognition of the merits of soft switching, with its promise of much improved performance at higher frequencies. One of the most interesting of these new developments is the thyristor-dual configuration, adapted as a phase leg in voltage-fed inverters. It appears that both MOSFETs and IGBTs are ideally suited for the synthesis of such an arrangement, in which the switches are controlled only at turn-off.

After reviewing the various commutation methods possible in converters, the attractions of soft switching are examined in detail, as well as the conditions necessary for its implementation. Power modules configured in the thyristor-dual mode are described, along with several application examples.

## **INTRODUCTION TO COMMUTATION METHODS**

First of all, it is necessary to distinguish between the manner in which a switch changes state and the external conditions causing this action. Stated differently, the switching action of the component itself must be segregated from commutation in the converter as a whole. Natural commutation, which appears to be optimal, is a natural phenomenon at converter level, where the spontaneous switch-off of a power semiconductor is provoked by the natural collapse of its current or voltage.

The study of commutation methods at switch level may be based on a few fundamental principles. Due to the essentially dissipative nature of a power semiconductor, it can only operate in the first and third quadrants of the voltage-current relationship.

A switching action resulting from the application of an electrical signal to a control electrode (the triggering of an SCR, for example), takes place in only one quadrant of the V-I plot, whereas a spontaneous switching action must be accompanied by a shift in operating point from one quadrant to another.

In order to minimize switching losses in the semiconductors, it is desirable that any changes in state be made along the V and I axes, or at least close to them. This is accomplished naturally with spontaneous commutation, but demands several precautions when forced switching is employed.

When one of the two switching actions in a semiconductor is spontaneous, the other being forced, the V-I characteristic must necessarily feature three parts, corresponding to either a current or voltage reversal. This type of switch is unique in its ability to be commutated automatically by the external circuit (example: an SCR possesses this property naturally).

Such components may be synthesized artificially, by juxtaposing together a device with controlled switch-on and switch-off properties, one of which is rendered automatic, along with the separate diode. A power MOS with its parasitic body diode, incidentally, provides this function without need for an external diode. The thyristor-dual, with controlled turn-off and spontaneous turn-on, is an example of a synthesized product (1),(2).

As a general rule with thyristor-duals, because only one transition is switched, it is relatively easy to limit the electrical stresses seen by the device during this switching interval.

**SOFT SWITCHING, CONDITIONS AND APPLICATIONS (10)**

Applying the preceding considerations, soft switching may be obtained when:

- The switch is defined by a V-I characteristic in three parts.
- There exists in the circuit a reversible source, capable of provoking spontaneous switch commutation at the right moments.
- Each power device is protected by a snubber circuit, to limit the stresses imposed across it during the one switched transition per cycle.

When these conditions are naturally filled in a converter, it is said to be naturally commutated. When the conditions are not satisfied, an auxiliary circuit must be added to circulate reactive energy (to force the current in a thyristor to zero before reapplication of forward blocking voltage, for example). Generally speaking, such auxiliary circuits consist of inductors and capacitors.

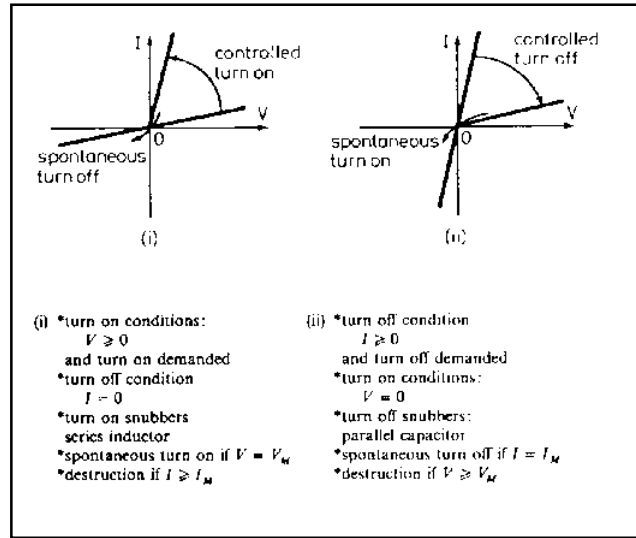
In practice, the need to discharge a snubber capacitor to zero volts (or a snubber inductor to zero current), introduces certain limits to converter operation; these limits define the characteristic operating areas of the converter being considered.

The scope of this paper is to analyze in detail the phase-leg topology associated with a voltage-fed inverter, where switched transitions occur at turn-off of the power semiconductors. This is a configuration for which MOS-gated power semiconductors are eminently suitable.

**THE INVERTER PHASE-LEG, WITH SWITCHED TURN-OFF**

In this topology, the switches must be thyristor-duals.

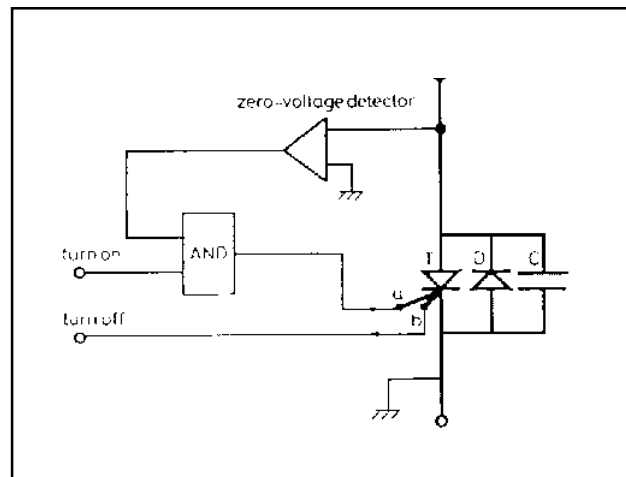
Figure 1 lists the principal characteristics of thyristor-duals, and as the name suggests, these are dual to those of conventional PNP thyristors (SCRs). In practice, thyristor duals are synthesized by connecting a semiconductor switch that can be turned off (BJT, MOSFET, IGBT, GTO or MCT) in antiparallel with a diode D and a snubber capacitor C (Figure 2).



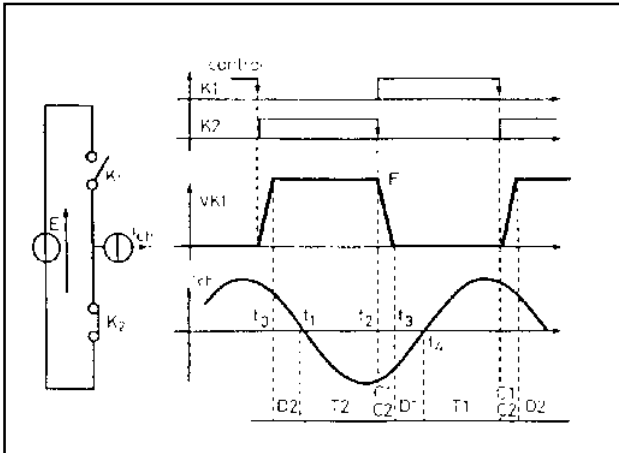
**Figure 1. Thyristor and Thyristor-dual**

In addition, they are also associated with a zero voltage detector that, in conjunction with appropriate logic, allows turn on only when zero voltage coincides with a turn-on control signal (AND gate).

The two switches, as well as all parameters and other elements associated with them, are differentiated in the schematics by the subscripts 1 and 2.



**Figure 2. Functional diagram of Thyristor-dual**



**Figure 3. Waveforms and control signals**

Referring to the waveforms of Figure 3,  $V_{K1}$  is the voltage across switch  $K_1$ ,  $I_{ch}$  the alternating current to the load, with  $K_1$  and  $K_2$  being gate signals to the two switches. For simplicity,  $I_{ch}$  is drawn here as a sinewave, but this is not restrictive, since only the position of its zero points with respect to the gate signals is important.

Starting from time  $t_0$  in Figure 3, the following operational sequences occur:

Just after  $t_0$ , the system state is defined by  $V_{K1}=E$  and  $V_{K2}=0$ ; corresponding to  $K_1$  being OFF and  $K_2$  ON. As long as  $I_{ch}$  is positive, diode  $D_2$  conducts. Because  $V_{K2}$  is zero, the turn-on control of  $T_2$  is active, so once  $I_{ch}$  changes sign (at  $t_1$ ), switch  $T_2$  takes over from  $D_2$  and continues to conduct until it is turned off. At  $T_2$  turn-off (at  $t_2$ ),  $I_{ch}$  starts to circulate in the two capacitors  $C_1$  and  $C_2$ , causing  $V_{K2}$  to rise and  $V_{K1}$  to fall. When  $V_{K1}$  has dropped to zero (at  $t_3$ ), diode  $D_1$  starts to conduct, and the turn-on control of  $T_1$  is activated. As soon as  $I_{ch}$  changes polarity once more (at  $t_4$ ),  $T_1$  can conduct until it too is turned off. The transfer  $T_1$ - $D_2$  then takes place in the same manner as  $T_2$ - $D_1$ .

**Fool Proof Operation:** It is important to realize that this operating mode prevents all chance of short-circuits in the branch. First, the turn-on signal of any one switch may only be activated when voltage across its mate is at  $E$ . This implies that the latter must be solidly off, generally with a negative gate voltage for maximum stability. Second, the taming of high  $dv/dt$  by the provision of snubbers allows the use, with a minimum of risk, of switches sensitive to this parameter. The bottom line is that this circuit topology is inherently safe, which also yields significant cost savings in the drive circuits, since the incorporation of dead time logic is not required.

**Diode commutation:** According to the operating mode described above, the power diodes in this configuration are commutated off at  $di/dt$  levels determined by the load current  $I_{ch}$ . These  $di/dt$  values are in general much lower than those found in hard switched topologies, where  $di/dt$  is usually limited only by lead inductance, snubber inductance, or transformer leakage inductance (in the case of classic SCR power supplies). Diode commutating voltage, moreover, is limited to the forward voltage drop of the companion power switches, a few volts at worst. In such conditions, problems routinely associated with diode reverse recovery just don't exist, to such a degree that it is even feasible in certain inverter modules to safely use the parasitic source-drain diodes of power MOSFETs.

**Commutation losses:** In all topologies employing turn-off semiconductors in association with snubber circuits, the blocking process is invariably accompanied by:

- Losses in the semiconductors themselves (non-zero voltage during current fall time).
- An exchange of energy between the load and the capacitors, since load current flows in the capacitors while voltage rises across the switches.

The method of evacuating this energy is, however, fundamentally different according to the principle of operation. In the case of DC choppers and classic PWM inverters, snubber energy is usually dissipated as heat during switch turn-on (RCD networks). In the case of the thyristor-dual inverter, though, this energy is dissipated in the load, since it is load current that is flowing in the capacitors during switch voltage collapse; all that occurs is a simple energy exchange between load and snubber. As a bonus, because the snubber is virtually loss-free, its capacitor value may be increased, thereby reducing semiconductor switch-off losses still further. In this configuration, turn-on losses are of course nonexistent (spontaneous turn-on).

It should be recalled that, in the case of RCD (or inductor-resistor) snubbers, energy dissipation occurs at switch turn-on (or turn-off). This imposes a minimum conduction time  $T_{on\ min}$  (or minimum nonconduction time  $T_{off\ min}$ ). Such a constraint does not exist for the thyristor-dual inverter, where only duration of the switching interval itself need be considered. In summary, this voltage-source inverter building block, with controlled switch turn-off, generates the lowest possible switching losses, while imposing no limitation on the minimum duration of  $T_{on}$  and  $T_{off}$ . It is, consequently, strategically placed for high efficiency power conversion at elevated frequencies.

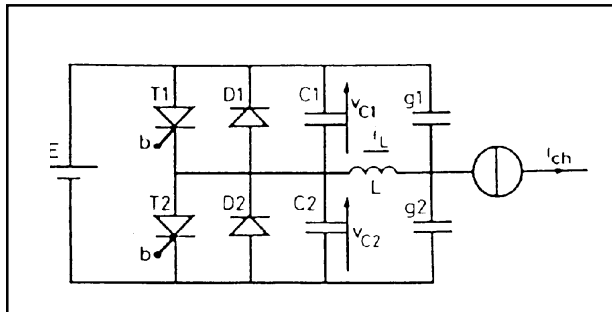
## CONVERTERS WITH SOFT COMMUTATION

Two different applications of the soft switched thyristor-dual building block will now be examined in depth.

### INVERTERS WITH A BI-DIRECTIONALIZED SOURCE (3)(8)

This name describes a family of inverter circuits that implement the chopper function with current reversal, or more generally, a PWM voltage source inverter function with soft switching. To accomplish this, various techniques are used to deliver alternating current from the voltage inverter at each switching transition, while maintaining unidirectional or low frequency AC current in the load. Under these conditions, the principle of a bi-directionalized source is created, along with the soft commutation that this engenders.

The functioning of such a converter relies on a sort of LC filter, interposed between inverter and load (Figure 4, (3), (12)). The elements of this filter are designed (11) such that current ripple in the inductance is always greater than twice the load current, and that the inherent frequency of the filter is less than the chopping frequency. Under these conditions, current delivered by the thyristor-dual inverter changes sign at each commutation point, thus fulfilling the requirements for soft switching.



**Figure 4. PWM inverter with filter and Thyristor-dual**

Operation as a sinusoidal inverter is naturally enhanced by the filtering intrinsic to this configuration. The low value of necessary filter inductance renders the filter transparent to the inverter fundamental frequency. In this way, control of the modulated voltage determines the output voltage, independently of the load, even when this latter is non-linear (a rectifier followed by a capacitor filter, for example (12)).

The naturally reversible nature of this circuit also favors its use as a rectifier with sinusoidal absorption characteristics (12). Waveforms are those of a PWM inverter, but with soft commutation and gently rising wave fronts.

In all variants, the necessity remains to circulate in the filter inductance a current at least twice that flowing in the load.

## RESONANT CONVERTERS

This configuration is based on the mating of a voltage source inverter featuring turn-off control, to a simple bridge rectifier. Because the bridge is unable to drive from the inverter suitably phase-shifted voltage current waveforms to realize soft switching, it is necessary to inject reactive energy via passive elements interposed between the two. The only control variable then possible is inverter frequency (non-modulated if soft switching is to be retained), but this renders the passive elements impedance variable.

Pure inductance is sufficient to store adequate energy for commutation of the controlled switch-off semiconductors, but the relationship between power dynamic range and frequency dynamic range is very limited. With a second order series or series-parallel circuit, selectivity is greatly improved. The commutation mechanisms depend on the relationship between the inverter switching frequency and the resonant circuit frequency, and are sometimes influenced by the load itself.

An example of this circuit is depicted in Figure 5(4).

Beyond the nonreversible form of DC-DC conversion, it is possible to envision other more sophisticated conversions based on resonance.

An alternating current controlled rectifier (5) (6) (inverse operation of the voltage-fed inverter) allows the realization of a DC-DC conversion, with reversible current capability both at the input and at the output. The series LC circuit guarantees impedance compatibility (voltage source/current source/voltage source) and ensures that, through judicious choice of working frequency with respect to LC circuit resonant frequency, sufficient reactive energy is available for both converters to function in the soft switching mode. Further out, it is possible to imagine DC to AC conversions for uninterruptible power supplies, or for AC machines, with unimaginable low levels of harmonic distortion (5) (6) (7) (9).

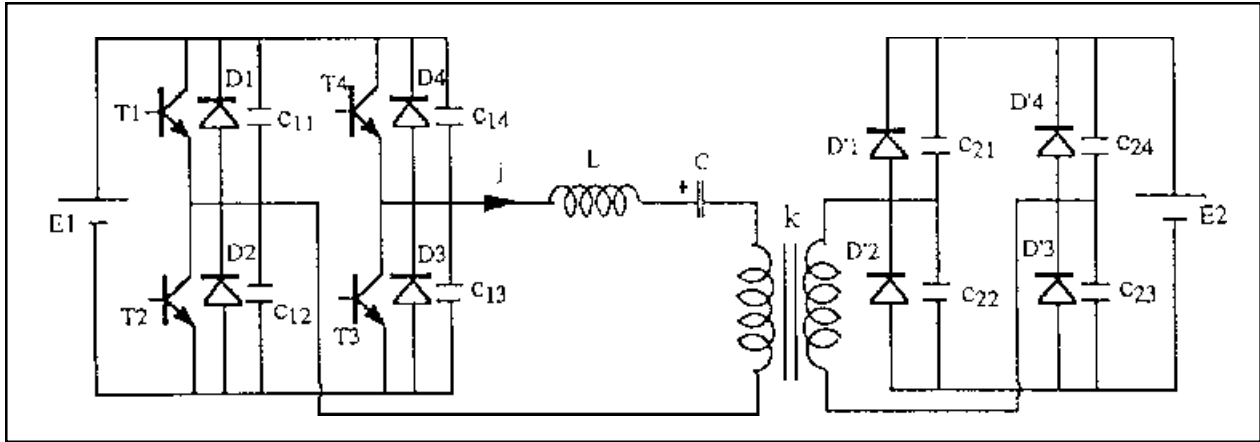


Figure 5. Series resonant converter

**WORKING EXAMPLES**

**PWM CONVERTERS WITH ZVS**

The ZVS/PWM topology can operate at efficiencies in excess of 90%, even when outputting several tens of KW at chopping frequencies above 20 KHz. The structure is well suited for the generation of either single or multi-phase AC power, when the end application requires a pure, distortion-free waveform (sinewave or other), with good regulation and fast speed of response. Such characteristics are particularly appealing when the load is nonlinear.

The main objectives targeted were:

- Minimum output voltage distortion
- Fast response time to sudden load changes
- Inaudible chopping frequency
- Minimum EMI
- Efficiency greater than 90%
- Compact modular structure

Thyristor dual modules LRGAT 75F100 and LRGAT 150F100 are optimized for use in isolated AC power supplies, working from three phase 440 VAC mains.

This power supply, rated at 15KVA, employs two Advanced Power Technology Europe (APTE) LRGAT 75F100 modules per phase. The configuration of one phase is portrayed in Figure 6. Development currently underway is aimed at boosting output to 90 KVA.

CIRTEM France, in collaboration with the University of Toulouse, has developed a 400 Hz AC power supply of this type, destined for the start-up and maintenance of naval airplanes on aircraft carriers.

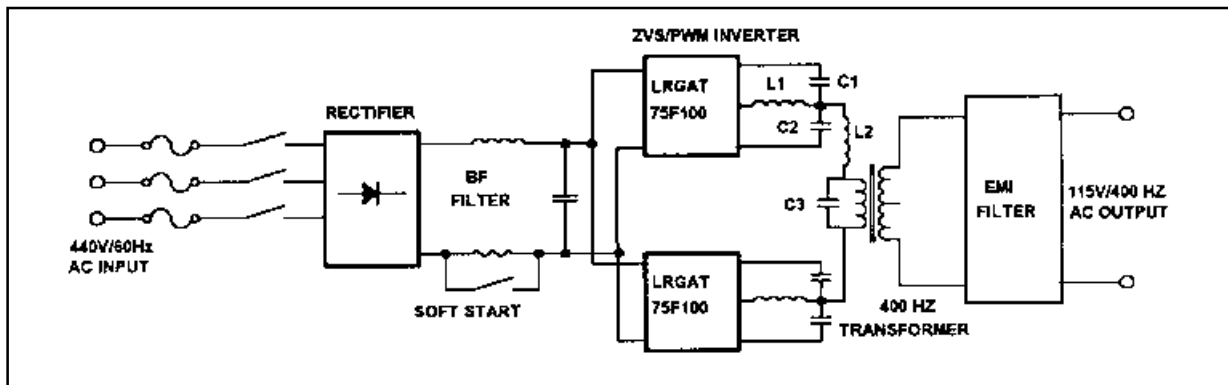


Figure 6. AC/AC PWM-ZVS Converter

An average chopping frequency of 25 KHz leads to such a reduction in filter size and insertion loss, that transient response time is less than 200us (16).

Minimum switching losses in the LRGAT 75F100 modules allow snubber reduction to 30 nF, thus ensuring safe switching operation well beyond the nominal output power limit.

Input Characteristics

- Nominal voltage: 440 VRMS +/- 15%, 60Hz three phase
- Power factor: > 90%
- Inrush current at power up: < Inom

Output Characteristics

- Nominal voltage: 115 VRMS 400 Hz three phase
- Harmonic distortion: < 4%
- Maximum power: 15KVA
- Efficiency: > 90% at nominal output
- Linear load: 0.95 leading < PF < 0.8 lagging
- Nonlinear load: according to Stanag 3456

Operating Characteristics

- Static voltage regulation: +/- 0.5%  
0 < P < 15KVA, any PF
- Static frequency regulation: +/- 2.5%  
0 < P < 15KVA, any PF
- Dynamic regulation: +/- 15% for deltaP +/- 0.33 Pnom

Experimental Results

Measurements made on a prototype 115V/400Hz/15KVA power unit are summarized below. Nominal DC bus voltage = 610V.

Figure 7 depicts current in L1 (20A/div, 500 us/div); this represents output current from one LRGAT 75F100 module. It consists of a 25 KHz (average) sawtooth, modulated at 400 Hz.

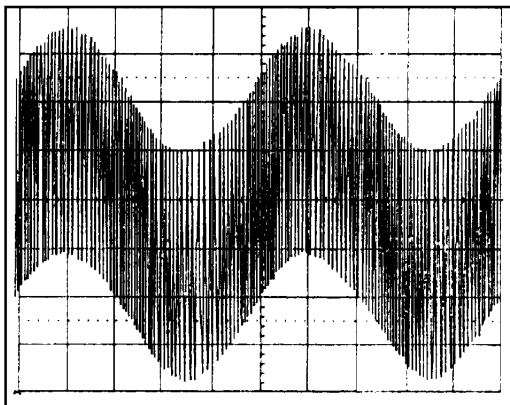


Figure 7. IL1 inductor current (20A/div, 500us/div)

Note particularly that the current changes polarity each period, a requirement for commutation of the LRGAT thyristor-dual module.

Figure 8 is an oscillation defining voltage across capacitor C1 (100V/div, 500 us/div). The fundamental frequency is 400 Hz, with HF ripple dependent on operating point; its average value is about 20%.

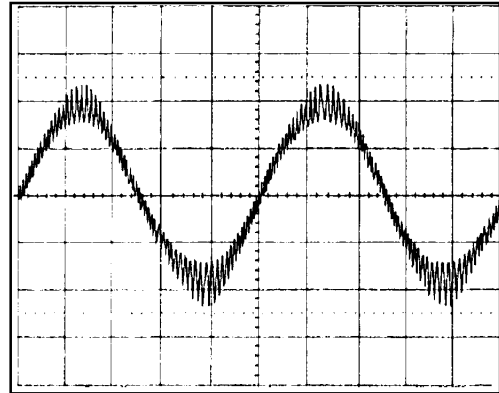


Figure 8: Voltage across C1 (100V/div, 500 us/div)

Figure 9 portrays output voltage and the current in a resistive load. Voltage across C1/C2 is first smoothed by L2C3, then stepped down by the 400 Hz output transformer. L2C3 attenuates HF so well that the output transformer sees virtually no ripple. Note the outstanding quality of the output waveforms; harmonic distortion < 1%. A visual comparison of choke current with output current illustrates the degree of current margin inherent to this topology.

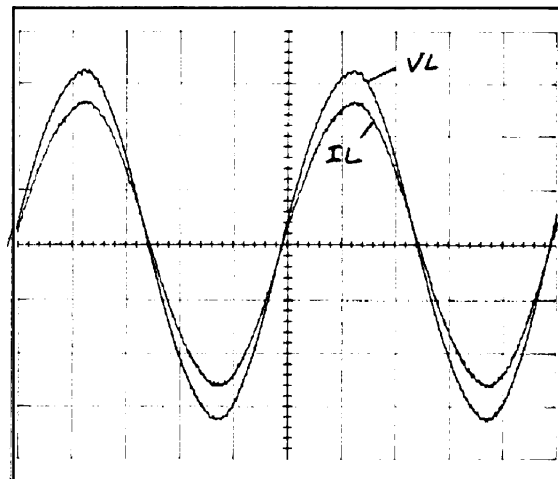
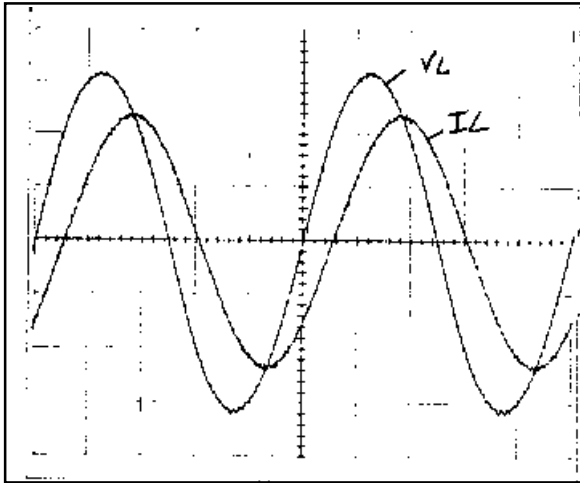


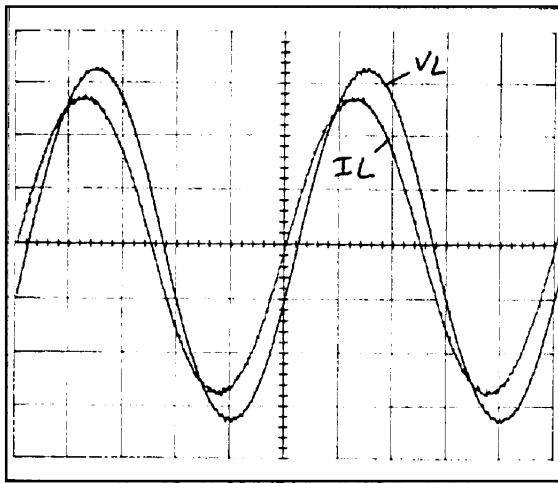
Figure 9. Voltage and current in the load (50V/div, 10A/div. 500us/div)

Figure 10 illustrates output voltage and current with an inductive load, PF=0.75 lagging.



**Figure 10. Voltage and current in the load**  
(50V/div, 10A/div, 500us/div)

Operation with a 0.95 leading PF capacitive load is shown in Figure 11. Both this oscillogram and that of Figure 10 attest to the purity of the output waveforms, even with reactive loads.



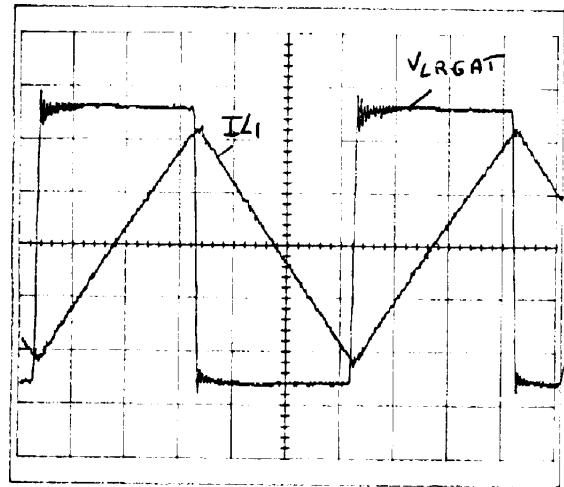
**Figure 11. Voltage and current in the load**  
(50V/div, 10A/div, 500us/div)

The final oscillogram of Figure 12 represents the voltage and current waveforms associated with the LRGAT 75F100 power switches, shown on a chopping frequency time base. Output current to the choke is triangular in shape, with no DC component.

This operational condition occurs at approximately the same time as the 400 Hz AC load current is crossing its zero axis. Within half period of chopping, there are two equal time intervals. The first interval, where IL1 is negative, corresponds to the conduction of an antiparallel diode connected across one of the IGBTs. The second interval, where IL1 is positive corresponds to the conduction of an IGBT.

Note the low value of dv/dt, about 1V/nS, characterizing the output voltage waveform; this is due essentially to presence of capacitive snubbers.

Thanks are due to H. Thiesen (15,16), who built this prototype, and undertook its evaluation.



**Figure 12. Module voltage and current**  
(100V/div, 20A/div, 5us/div)

**SERIES RESONANT CONVERTER**

The series resonant DC-DC converter, described in articles (5, 10), is particularly suitable for applications where substantial output power (tens of kW) must be generated at high efficiencies (> 90%), with chopping frequencies above the audio spectrum (f > 20 kHz). It is appropriate to note that this technology is very well adapted to high voltage DC power generation, due to the favorable commutation conditions existing when the resonant inverter is married to a secondary rectifier.

The LRGAT75F100 and LRGAT150F100 thyristor-dual modules described earlier are equally suitable for use in series-resonant power conversion, especially when input power is sourced from 400 VAC three phase mains, and the output must be isolated. A variety of power supplies based on this technology have already been developed (for laser drivers, magnetrons, power tubes, etc.). One such equipment, a 30 KW/ 120 VDC power supply, will be examined in detail.



### 30KW/120VDC Power Supply

To fill a need to test energy conversion groups, CIRTEM has produced several of these 30 KW/120VDC power supplies. The key design objectives were:

- Minimum size and weight (l = 19", h = 12U; M < 100Kg @ 30KW)
- Inaudible chopping frequency
- Minimum output ripple
- Efficiency > 90%
- Compact and modular structure, to fit standard 19" rack

A series resonant converter was chosen as being best able to satisfy the design brief.

A block diagram of the power supply is portrayed in Figure 13. It uses two LRGAT150F100 thyristor-dual modules; switched current is in the region of 100A at 25 kHz. Power output is varied by changing chopping frequency.

Its principal features are:

#### Input Characteristics

- Mains voltage: 400VRMS +/- 10%, 50/60 Hz, three phase
- Power factor: > 0.9
- Inrush current at start-up: < Inom

#### Output Characteristics

- Variable output voltage, in two ranges:  
Parallel connected: 0 - 120VDC at 250A  
Series connected: 0 - 240VDC at 125A
- Maximum output current adjustable on each range
- Maximum power: 30KW
- Efficiency: > 90% at nominal power

#### Operating Characteristics

- Static regulation: +/- 1%, assuming cumulative load and mains variations, load 5 to 100%, mains +/- 10%
- Electronic output current limit at:  
312A when parallel connected  
156A when series connected
- Ripple content: +/- 1% peak to peak

#### Operating Conditions

- Ambient temperature 0 to 40°C
- Forced convection cooling

### **THE LRGAT POWER MODULES**

APTE, in collaboration with the CIRTEM company and Toulouse University, has engineered a range of standard power modules, configured as phase legs (half bridges) to operate in the ZVS/PWM thyristor-dual mode. Featuring 1000V rated IGBTs with matching antiparallel diodes, these modules also incorporate optimized clamping networks and logic circuitry to provide full ZVS functionality. Galvanically isolated driver circuits are positioned in close proximity to the power switches for best performance.

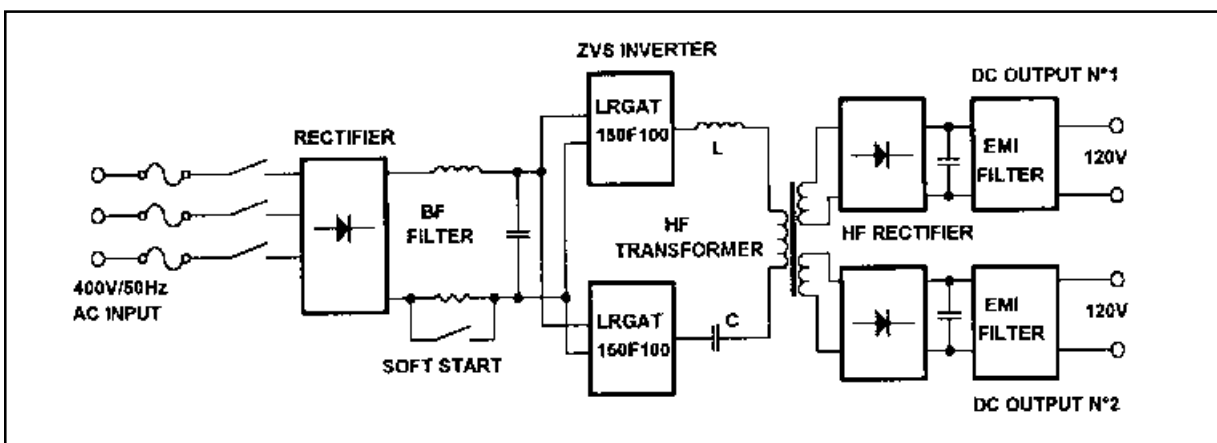


Figure 13. 30KW/120V resonant power supply

Control signals are isolated via high frequency transformers, which enhance electrical performance as well as reliability.

These transformers also transmit drive power for the IGBTs, thereby obviating the need for auxiliary power supplies on the secondary side. A single 15V supply on the primary side is all that is required for full compatibility with CMOS control logic.

Start-up circuitry ensures that the system will power-up properly when full voltage is applied, by temporary inhibition of the ZVS mode for a few microseconds.

The module block diagram is illustrated in Figure 14.

Product is available in 75A and 150A versions, both housed in standard LP8-outline packages. Although this low-profile package is extremely compact, it is good for output power levels up to several tens of Kilowatts (see Figure 15).

While the efficiency of ZVS converters of this type is unmatched at high frequencies, power dissipation in the module is still significant, given its compact nature. In order to minimize external cooling system size and cost, yet keeping junction temperature modest for longest life, the power semiconductor chips are mounted on aluminium nitride substrates for best thermal conductivity.

Driver output stages, implemented with both SMD and chip components, are mounted on the same ceramic substrate as the power devices. Control and isolation circuits, on the other hand, are mounted onto a four layer PCB, one layer of which is a ground plane for best noise immunity.

Power connections are made to the modules via M5 screws, whereas logic level and auxiliary supply circuits are interfaced through 0.6 mm x 0.6 mm pins on 2.54 mm centers; this arrangement permits direct mounting of a control board without the need for wire links. In this way, parasitics are minimized, and reproducibility is assured over very long production runs.

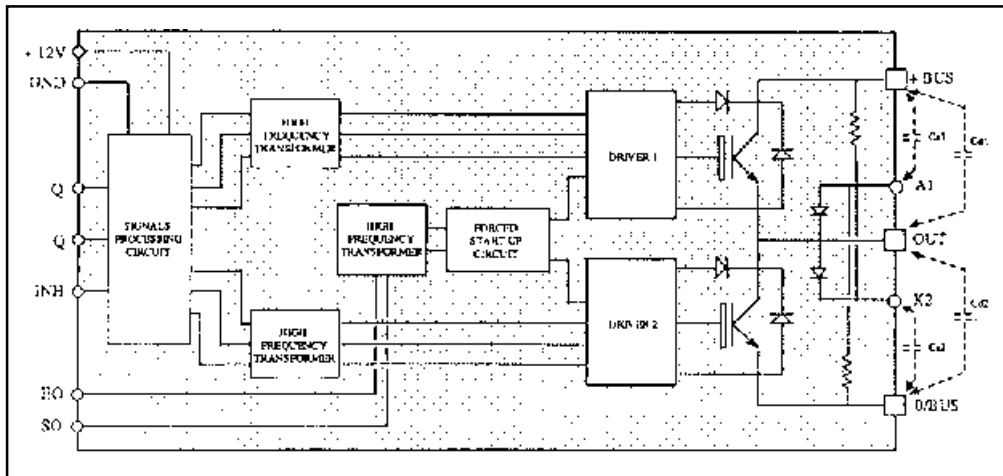


Figure 14. LRGAT module block diagram

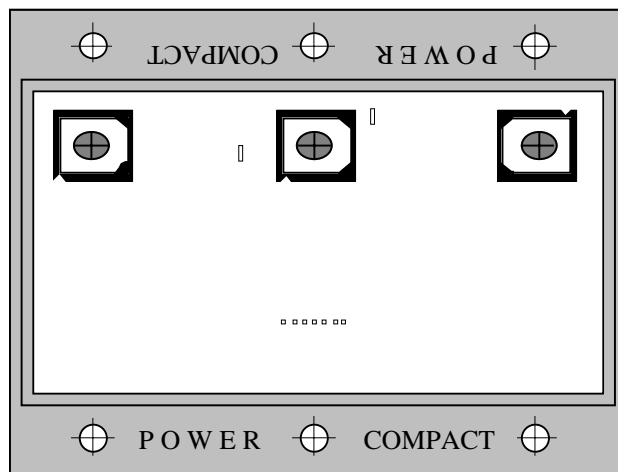


Figure 15. LRGAT module pinout location

## MATCHING MODULE TO LOAD

The two examples that follow exploit to the full the power handling capabilities of LRGAT150F100 modules, in both AZS/PWM and series resonant converter applications.

The AVS/PWM inverter of Figure 16 uses a single half bridge module, with a capacitive middle point to complete the bridge. This inverter outputs 5 kVA while operating at a maximum frequency of 40 kHz. Peak switch current is 134A.

The series resonant DC power supply, illustrated in Figure 17, again uses a single LRGAT150F100 module with a capacitive voltage divider to complete the power mesh.

Output power here is 16 KW, the nominal operating point corresponding to 70A at 230VDC. Switching frequency is 23 kHz.

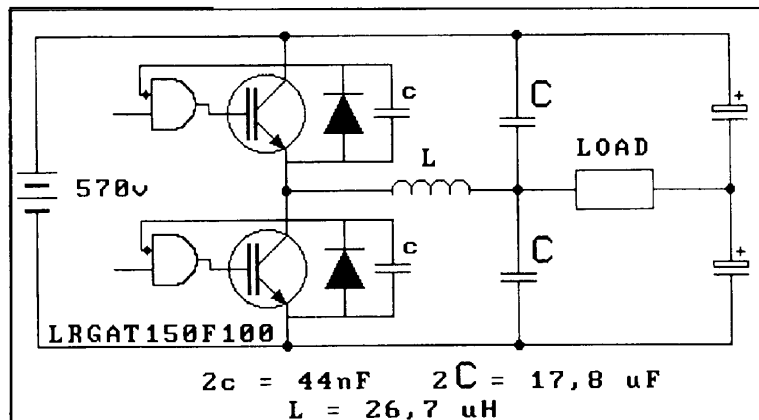


Figure 16. PWM-AVS converter

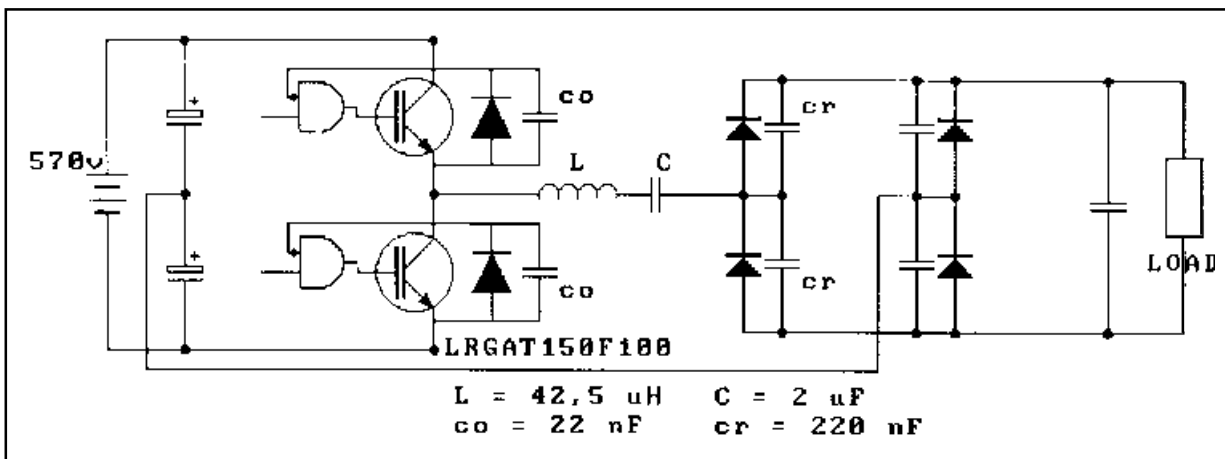


Figure 17. Series resonant power supply

Figure 18 highlights several key voltage and current waveforms under various operating conditions,

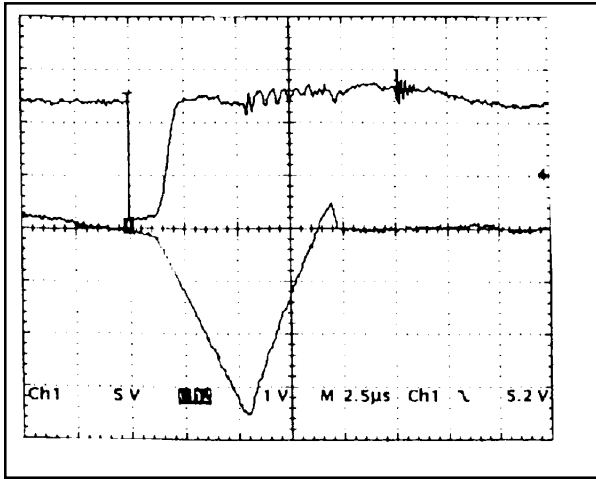


Figure 18a. 300A/500V turn off

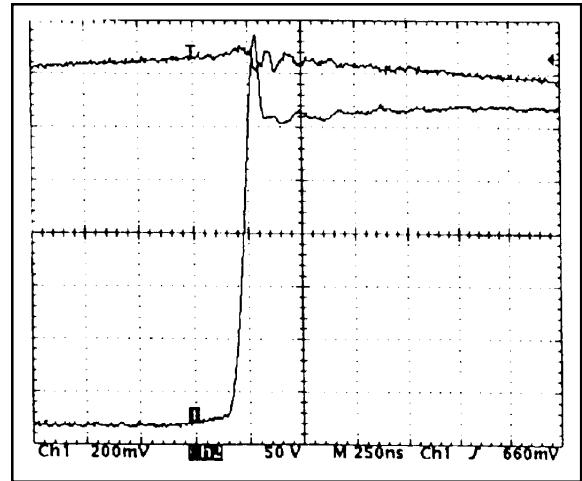


Figure 18b. 50A/300V turn off without snubber

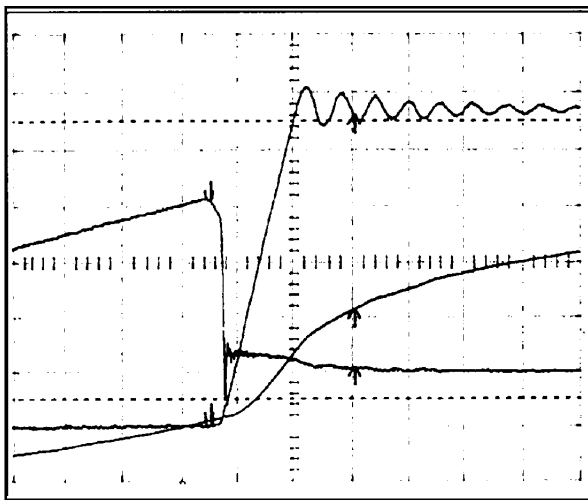


Figure 18c. 60A/550V turn off at 25°C

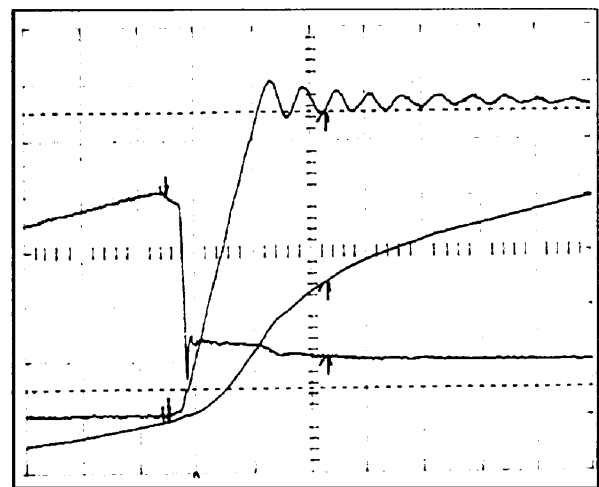


Figure 18d. 60A/550V turn off at 100°C

## CONCLUSION

Thanks to the inherent compactness and reliability of hybrid technology, these power modules facilitate the construction of state-of-the-art converters, both AC and DC, capable of outstanding and uniform performance. The LRGAT module range that has been described, being based on IGBT technology, is most suited for very high power applications operating from 400VAC mains. The IGBT technology, nonetheless, limits chopping frequencies to the low tens of kilohertz.

By specifying power MOSFETs, instead of IGBTs, APTE, working with the same partners, has now developed resonant power modules optimized for 230 VAC applications. These power MOSFET based modules are capable of switching at frequencies up to 400 kHz, and are ideal for very high performance low volume converters.

Finally, these ZVS/PWM modules, whether IGBT or power MOSFET based, represent an optimum solution for a wide range of output power, voltage and frequency requirements. Easy to use, thanks to a high level of integration, they allow the equipment designer to engineer simply constructed high performance converters using only power module building blocks, together with a few external passive components and minimum support circuitry.

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