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PROBLEMS, SOLUTIONS AND APPLICATIONS IN THE DEVELOPMENT OF A WIDE BAND POWER AMPLIFIER FOR MAGNETIC BEARINGS

T.BARDAS*, T.HARRIS*, C.OLEKSUK*, G.EISENBART**, J.GEERLIGS**

*NOVA CORPORATION OF ALBERTA
**NOVA HUSKY RESEARCH CORPORATION

Abstract

This paper describes the development of a power amplifier for magnetic bearings capable of switching 50 A and 160 V DC at 40 kHz. Magnetic bearing amplifiers for large industrial equipment require high power handling capacity to control bearing current. In order to reduce power losses, the amplifiers need to be switching. A high clock rate in the amplifier is desirable to reduce distortion in the bearing current but causes increases in the amplifier power losses. The operation of a switching amplifier is described as well as the limitations of the components. The testing results of the amplifier are discussed.

1. Introduction

NOVA Corporation of Alberta currently operates 13 gas compressors ranging in size from 3,800 kW to 21,000 kW that utilize magnetic bearings. The magnetic bearings used in NOVA's compressors have been supplied by Magnetic Bearings Incorporated and operate using an attractive principle. The current in the bearings is used to create a magnetic field between the rotor and stator elements. The current and, therefore, the bearing force is controlled by a switching regulator commonly called a power amplifier.

With an eye towards the future and machines with more flexible rotors operating at higher speeds, NOVA saw the need for a more efficient and effective amplifier. NOVA, therefore, commissioned the development of such an amplifier to its research arm NOVA HUSKY Research Corporation. The amplifier was to have the same power output as the present amplifiers (50 A x 160 V) but the switching speed was to be more than doubled from 17 Khz to 40 Khz. The amplifier cooling was to be forced ambient air only.

2. Power Output Requirements for Power Amplifiers

An important characteristic in determining whether an active magnetic bearing can modulate a dynamic load is the rate at which force can be changed in the bearing. Assuming dynamic forces on a shaft to be sinusoidal in nature, higher force change rates result from larger forces or higher frequencies. If the actual rate of force change exceeds the bearing's maximum force change rate, the bearing will be unable to control the load and is, therefore, force slew rate limited. The sum of the dynamic and static forces on the bearing may be below the bearing's static capacity but the bearing current cannot be changed fast enough to access the unused capacity. On an actual machine, the current wave form becomes triangular in nature when the bearing is force slew rate limited.

The parameters that control the rate of force change in a bearing, can be found by multiplying the partial derivative of force with respect to current, by the derivative of current with respect to time. (see Appendix 'A').

$$\frac{dF}{dt} = \frac{\partial F}{\partial i_b}\frac{di_b}{dt} = \frac{i_b V_b}{g} \qquad (1)$$

The maximum rate of force change in a bearing is governed by the gap in the bearing and by the power applied to the bearing by the power amplifier and is unrelated to the size of the bearing. In practice reducing the air gap does not change the bearing force response because halving the air gap in a given radial bearing also halves the static current needed to levitate the shaft. The only methods to improve the bearings force response are to increase the voltage driving the coil or the static operating current. The static operating current can be increased by reducing the number of turns or by increasing the static load with an opposing magnetic bearing. In either case the amplifier must be able to deliver the necessary current at the required voltage in order to make the bearing response satisfactory.

In NOVA's compressor stations the D.C. back-up power for the station is provided by stationary batteries that float at 132 V. The cost and space requirements for a separate battery bank outweigh the benefits of operating the bearing at a higher voltage. The air gap in the radial bearing is usually about $5.1x10^{-4}$m while the levitating current is about 20 amps. From equation (1) this results in a maximum rate of force change of $5.2x10^{6}$ N/sec.

The maximum actual rate of force change for an unbalance is:

$$\frac{dF}{dt} = \frac{2\pi n F_{peak}}{60} \quad (2)$$

Therefore, a typical radial bearing operating at 10,000 rpm could produce the force change rate developed from a 4,945 N peak unbalance force. This only becomes a limitation if the bearing has capacity for unbalance loads greater than 4,945 N peak.

3. Switching vs. Linear Amplifiers

In order to control the field in a magnetic bearing, regulation of current to the bearing by an amplifier is necessary. The simplest choice for current control is a linear amplifier consisting of a transistor in series with the bearing coil as shown in Fig. 1. The bearing current is varied by changing the voltage drop across the regulating transistor. A current feedback loop is used to regulate the voltage drop across the transistor. This approach has been used successfully in magnetic bearings but leads to high heat dissipation in the amplifier when large static or dynamic load capacities are required.

The power needed by a linear amplifier is determined by the supply voltage and the current in the bearing.

$$P_q = V_s i_b = i_b(V_t + V_b) \quad (3)$$

The heat dissipated in a linear amplifier is determined by the voltage drop across the transistor and the current. The ability of the bearing, to respond to force change is determined by the additional voltage above steady state that can be placed across the bearing coil. The additional voltage that can be placed across the coil is the voltage drop across the transistor. Therefore the ability of the bearing to regulate dynamic loads is directly related to the power losses of the amplifier.

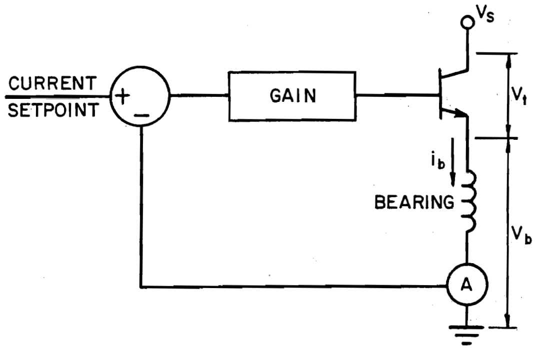

FIGURE 1
LINEAR POWER AMPLIFIER

As an example, a typical quadrant in a radial bearing used in a gas compressor requires 20 amps to levitate the shaft. The amplifier is powered by 130 V and the coil resistance is about 0.3 ohm. The steady state voltage drop across the coil would be 6 V and the drop across the transistor would be 124 V. The power required for this axis, in levitating the shaft, would be 2,600 W of which 2,480 W is lost in the power amplifier.

A compressor normally has 4 active radial bearing quadrants which would result in nearly 10 kW of power being required for amplifier losses. While the 10 kW may be insignificant in terms of energy compared to the compressor, it represents a significant problem for cooling the amplifiers to a temperature at which the regulating transistors can survive.

The alternative to a linear amplifier is a switching amplifier. The simplest possibility is as shown in Fig. 2. When the current required is below set point the switch is connected to the voltage supply. The current would then rise at a rate of

V_b/L_b until it reached the desired level, when the switch would close shorting the coil. The voltage across the coil would reverse and the current would continue. The rate of current decay would be determined by the L/R time constant of the coil and the bearing current. Normally, this would result in an unacceptable length of time to reduce to the current and the bearing force due to the low resistance in the loop.

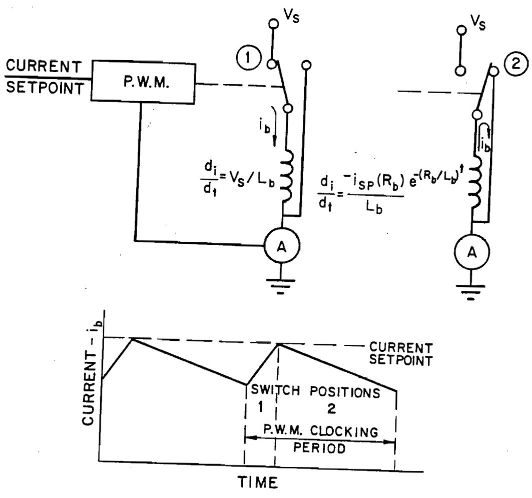

FIGURE 2
SINGLE SWITCH SWITCHING AMPLIFIER

An additional resistor could be added in the closing loop to decrease the L/R time constant to make the rate of current decay equal to the current rise for a given operating current. This results in the resistor being sized to produce the same voltage drop as the voltage supply. With equal rise and decay rates a duty cycle of 50% is required in order to maintain the selected operating current. Half of the time the bearing receives energy from the power supply, the other half the bearing supplies energy to the flyback resistor. The bearing axis losses would then be:

$$P_q = i_b^2 R_b + \frac{V_b i_b}{2} \qquad (4)$$

The power losses in a bearing quadrant using the previous example would be 1420 W with 1300 W being lost in the flyback resistor. The amplifier loses are still high and the current decay rate varies with the bearing current.

The ideal situation is if the energy lost in the fly back resistor could be fed back in to the power supply. The energy could then be reused in the next switch cycle when the current in the bearing is being increased. This can be accomplished by a circuit shown in Fig 3. The supply voltage is applied when switches 1 and 4 are closed and switches 2 and 3 are opened. The current increases at a rate of V_b/L_b. When switches 2 and 3 are closed and switches 1 and 4 are opened, the voltage across the coil reverses and the current flows from the coil back to the power supply. The current decreases at rate of V_b/L_b. The power supply must be able to accept the current from the bearing or high coil voltage will result. A capacitor across the power supply serves this purpose. The amplifier is really a switch that transfers energy back and forth between an inductor and a capacitor.

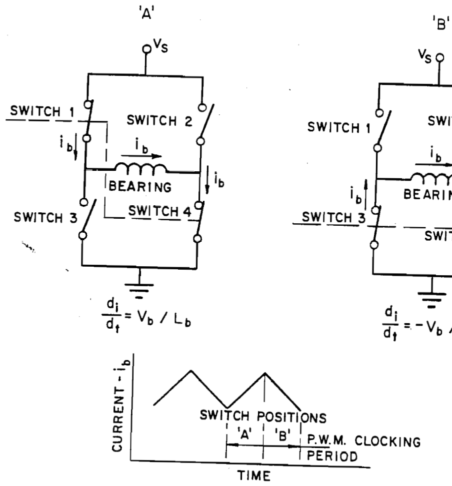

FIGURE 3
FOUR SWITCH SWITCHING AMPLIFIER

The drawback of using a switching amplifier is the distortion introduced into the bearing current. At a constant current level the peak to peak value of the ripple current is given by the following:

$$i_r = \frac{V_b}{2L_b f_{pwm}} \qquad (5)$$

The amount of ripple, therefore, depends on the size of bearing and the switching frequency. In small machines the switching frequency must be increased to reduce the ripple to an acceptable amount.

The amplifier losses are restricted to resistive losses across the switches and the losses during switching. By proper design these losses can be made a fraction of the product of voltage and current regulated by the amplifier. This is the design challenge.

4. Operation of a Switching Amplifier

The switching amplifier (Fig. 4) consists of two transistor switches and two diodes arranged to form an H-bridge with the bearing coil. The two transistors form one pair of switches that allow energy to be increased in the bearing. The diodes form the other pair of switches that allow energy to be reduced in the bearing. The transistors are the only switches that have to be actively controlled. The diodes are passive switches that become conducting whenever the coil voltage becomes greater than the supply voltage. This guarantees the current direction in the bearing and that the coil will not become open circuited. The high resulting voltages could destroy components.

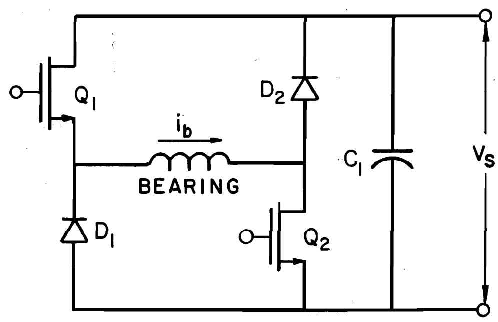

FIGURE 4

SWITCHING AMPLIFIER SCHEMATIC

The switching of the amplifier is controlled by a pulse width modulator. The pulse width modulator (PWM) can be controlled using either current or flux feedback. The pulse frequency of the PWM, therefore, controls the maximum switching rate of the amplifier. During operation at any constant current, the amplifier duty cycle is nearly 50%. When the coil current is increasing the duty cycle increases.

Regulation of the flux density in the bearing is achieved by switching energy between the capacitor and the bearing coil. This results in the highly efficient nature of switching amplifiers, as compared to linear amplifier which regulates by dissipating energy in the form of heat.

The capacitor sees an alternating current equal to the switching frequency of the amplifier. The peak value of the alternating current, equals the bearing current. The inductance of the power supply cable prevents most of the switching current from being seen at the power supply. The current from the power supply is mainly D.C. and is determined by the power losses in the bearing and amplifier.

5. Power Losses in Switching Amplifiers

The power losses in the amplifier can be divided into two categories, static and dynamic. The static losses are caused by the resistance of the transistors and the capacitor and the forward biased voltage drop of the diodes. These losses are related to the square of the bearing current in the case of resistance and directly to the current in the diodes. The static power losses can be minimized by choosing metal oxide semi-conductor field effect transistors (MOSFET) with low resistances. In the case of the diodes the voltage drop is fixed if silicon diodes are used.

The dynamic losses result from the fact that the transistor switches do not start and stop conducting current instantaneously. When one of the transistor switches turns on, the full supply voltage is across the device. The voltage remains until the transistor resistance is low enough to allow it to conduct the full bearing current at which time the the voltage drop is determined by the transistor resistance. A similar situation occurs when the transistor turns off. The dynamic losses are related to the bearing current and the switching frequency of the amplifier. In order to minimize the switching losses in the amplifier, the transistor switches must be selected to switch as quickly as possible. As a result MOSFET transistors were selected.

6. Limitation of Amplifier Components

According to initial calculations, power losses of less than 150 W were attainable. Actual measurements of a breadboarded switch, however, proved that dynamic losses were greater than expected. In order to keep these losses low at the 40 kHz switching rate, the switching time had to be kept to a minimum. Unfortunately very little information was available about high speed, high voltage and high current switching problems and the initial tests were not encouraging due to high heat dissipation. The development of the switch was further aggravated by severe oscillations which could not be suppressed by means of dampening. Contrary to the expectations of high efficiency when using fast recovery diodes and MOSFET switches, the limitations of such devices and their undesirable interactions were observed. These are explained in connection with problems encountered during the development.

The immunity of MOSFET's to thermal second-breakdown failures gives them an advantage over bipolar transistors. Yet they still run the risk of avalanche breakdown as do bipolar devices. The avalanche breakdown can be traced to the turning on and off of the parasitic bipolar transistor which is an inherent feature of each MOSFET device (Fig. 5). The voltage limit of the parasitic transistor is its base-collector junction breakdown voltage. Its base-emitter resistance, temperature and D.C. current gain (h_{FE}) determine the current level at which the breakdown occurs. In the MOSFET's parasitic transistor, the base and emitter terminals are shorted together on the die, to yield the least base-emitter resistance (R_{BE}). The base to emitter resistance together with the collector-base junction capacitance (C_{CB}) determine the MOSFET's drain to source dV/dt. A positive transition of the drain to source voltage causes flow of current through the C_{CB} which corresponds to the rate of the voltage rise. The resulting voltage drop across the R_{BE} can reach a point, at which the parasitic transistor turns on. This undesirable turn on or switchback, interferes with the normal circuit operation of the MOSFET and can possibly destroy it.

Maximum dV_{DS}/dt information is not yet specified on MOSFET data sheets. The high rate of voltage rise between the drain and the source of the MOSFET can only occur during the off-

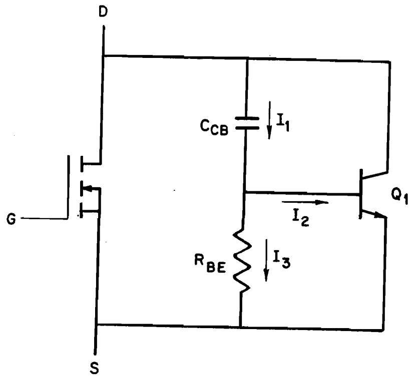

FIGURE 5
EQUIVALENT MOSFET SWITCH SCHEMATIC WITH THE INHERENT BIPOLAR TRANSISTOR

switching of an inductive load. Such is not the case with the H-bridge switch, since the fast recovery rectifiers effectively clamp the bearing magnet to the supply. In spite of that, some parasitic inductances associated with the H-bridge switch cause severe voltage spikes. In order to reduce the effects of parasitic inductances, interconnections must be as short as possible within the switch.

Data sheets of the MOSFET used, however, list a maximum allowed rate of change of current at turn off (di/dt). The value of di/dt depends on the drain to source voltage. For the 160 V supply voltage, a maximum of 10 A/ns can be read. Actual measurements have shown a rate of 1 A/ns under maximum load conditions. The maximum rate of current change might have drastically increased during critical conditions outlined in the test result section.

Another limiting factor of the MOSFET is its breakdown voltage limit. As the breakdown voltage increases, the transistor's on-resistance goes up exponentially. In addition to the supply voltage a reasonable safety margin must be added for the handling of voltage surges. For the 160 V supply, a 400 V MOSFET device was considered adequate.

The switching speed of a MOSFET device is determined by the travel time of the majority carriers, the gate capacitance, the transistor's package inductance, the internal gate resistance and the driver impedance. The delay at turn on is due to the length of time it takes for the gate voltage to rise to the threshold level, where the device begins to conduct. As the gate to source voltage increases above the threshold level, the drain to source voltage is falling and the Miller effect takes place while a large increase in capacitance occurs. After the drain voltage drops to its final "on value", the gate to source voltage begins to increase again and no Miller effect takes place. A typical characteristic of the drain-source voltage versus the gate charge is shown in Fig. 6. Particularly critical is the time during the drain to source voltage drop during which severe oscillations were observed. A quick charging of the gate through a low impedance drive circuit seemed to be an easy solution. Unfortunately, the MOSFET power module has no internal Zener diode protection and the maximum gate to source voltage is only 20 V. It is, therefore, important to select devices with short duration of drain to source voltage drop.

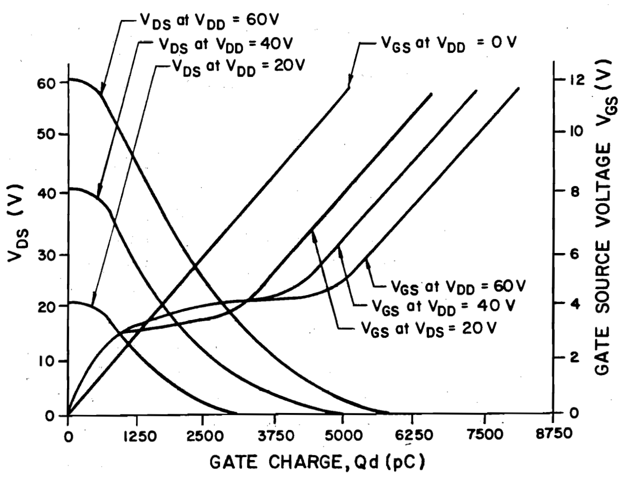

FIGURE 6

DRAIN-SOURCE VOLTAGE VERSUS GATE CHARGE

The reverse recovery time (t_{rr}) of the 4-bridge rectifier has been singled out as the main cause of the circuit oscillations. During the reverse recovery time, diodes continue to conduct in the reversed voltage direction. The current is determined by only the external resistance in the diode circuit. The reverse recovery time should be as short as possible because during that period, both rectifiers of the switch represent a short in the respective leg of the H-bridge. The fast recovery rectifiers used in the bridge have a trr of 50 n sec. The mechanism of circuit oscillations can be explained by means of the H-bridge diagram into which parasitic capacitances and stray inductances have been added as shown in Fig. 7. In the diagram C_s represents the capacitance associated with the rectifier. Its value is rather variable and dependent on the state of the device. In this instance it is considered to be in the "off" state. L_S is the stray inductance representing the wiring and the lead inductance of the rectifier. Cgd is the gate to drain capacitance of the MOSFET and it too is variable according to the state of the device. To simplify the discussion, the other stray inductances and parasitic capacitances at the various MOSFET leads are not shown. A series resonant circuit is formed by C_S and L_S in each leg of the bridge. Whenever it is excited by a pulse, the circuit will ring at its natural frequency. Without further intervention, the decay time of the oscillations can be many microseconds. At the MOSFET turn off, this oscillation barely occurs because the diodes switch to their "on" state and C_S is effectively removed thus preventing the energy exchange with L_S. At the MOSFET turn on, however, the effect of the resonant circuit is dramatic and potentially destructive. Within less than 100 nanoseconds the MOSFET carries its full rated current and the

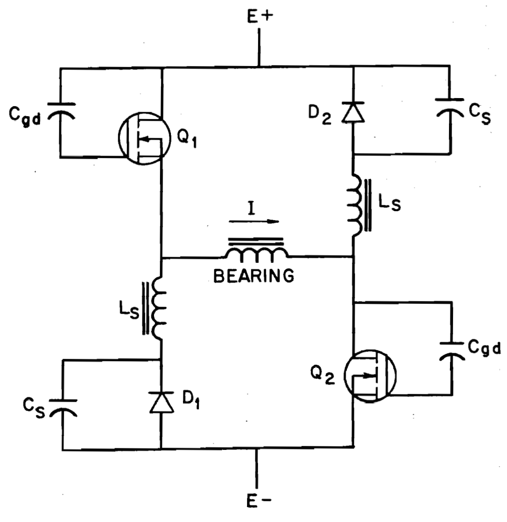

FIGURE 7

H-BRIDGE WITH STRAY COMPONENTS

rectifier begins its turn off. At this time, the drain voltage is still high and within 50 nanoseconds after the load current has been taken up by the MOSFET, the reverse recovery charge of the rectifier collapses. The energy in the stray inductance is now exchanged with the parasitic capacitance C_S in an oscillatory fashion at frequencies of 60 to 150 megahertz. With little damping, very large voltage oscillations are present at the drain which is coupled into the gate through the gate to drain capacitance Cgd. The net result is increased power dissipation in the MOSFET as well as large voltage swings at the gate which can destroy the device.

Thus our concerns were concentrated on the basic H-bridge circuit. The problems were further aggravated by circuits external to the power amplifier. The power supply by necessity is some distance away from the amplifier and is connected with long cable runs. These cables have inherent inductances, that are rather large, in the hundreds of microhenries range. As far as the switching problems in the bridge are concerned, the net result is the same as was caused by the stray inductances inside the bridge but on a larger scale. In addition, since the full load current reverses polarity every half cycle, these cables would have to be inordinately large if more than one amplifier was in use. For this reason, an electrolytic capacitor bypasses the supply bus bar in each amplifier. Unfortunately, electrolytic capacitors for higher voltage ratings have poor characteristics in regard to the series inductance (ESL) and the series resistance (ESR). The ESL figure limits the applicability of electrolytic capacitors for high di/dt demands, whereas the ESR value determines ohmic losses at a given ripple current.

These are all problems due to inherent characteristics of components. Solution of these becomes the key to a successful design.

Measurements taken during the laboratory tests of the amplifier is recorded on diagrams shown in Fig. 8. The oscillograms of the output voltages, which were measured between the load terminals and the ground, verified the switching speed. The barely visible voltage spikes are within a safe operating limit of the semiconductor components.

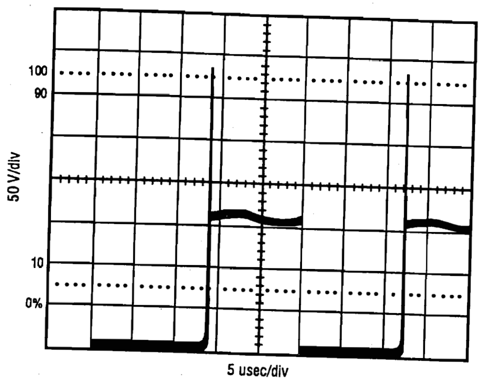

FIGURE 8
Amplifier Load Terminal 1 to Ground Voltage

7. Design and Layout of the Power Amplifier

Performance criteria of the power amplifier was achieved by a careful selection of components as well as efficient layout and mounting of all parts. The design goal was for a high speed H-bridge switch that could be operated with air cooling. Because at least ten amplifiers are used in the typical magnetic bearing installations, physical size considerations were important. Cabinet space is always at a premium. Fortunately, compact size coincides with the electrical considerations to reduce stray inductance through minimal conductor lengths.

Once the appropriate heat sink was selected, all the electrical components were fitted to the mounting surface of the heat sink. Efficient MOSFET's were selected and mounted directly on the heat sink, then interconnected by heavy bus bars. For ease of mounting, power modules were preferred to flat pack housing. Rectifier diodes are mounted to the heat sink on a heavy L bracket. A printed circuit board, carrying all remaining components is positioned on top of the bus bars. The arrangement features short, low inductance paths among all components. An example of optimization is the connection between the gate driver chips and gates of the power MOSFET's. In this instance, an excess length of only 10 mm would cause a 60% increase in spurious noise on the gate signal in spite of the fact that the drivers can source or sink 6 A.

Effective dampening of oscillations has been achieved by RC snubbers as well as losses of ferrite cores which are coupled to the conductors of the bridge. Dampening by ferrite has some negative effect on the stray inductances which create voltage spikes, so care must be made in the application.

A pulse width modulated signal at the output of the controller is sent onto a fibre optic cable through a fibre optic transmitter. The cable carries the signal to a fibre optic receiver at the amplifier and appropriate drivers control the gates of the MOSFET switches. The fibre optic signal transmission avoids any potential noise pick-up due to EMI effects caused by high dv/dt or due to the number of amplifiers in use.

8. Amplifier Performance Results

The original specifications for the development of the amplifier called for operation at a 40 kHz switching rate with the power dissipation of less than 150 W at a maximum load of 50 A and 160 VDC. The switching rate and load capability were fully satisfied. The heat dissipation spec was not met.

At the present time, available power measurement equipment revealed a power loss of approximately 200 W at a room temperature of 21°C. The corresponding efficiency amounts to 97.5%. An increase of efficiency is expected at a lower switching rate of 17 kHz, however, it was not yet confirmed by measurements. Although heat dissipation is higher than specified the amplifier is still suitable for air cooled cabinet design.

Ruggedness of the amplifier has been proven by breaker trips under short circuit conditions, which did not cause any deterioration of performance. The design is, therefore, considered suitable for industrial service.

Measurements taken during the laboratory tests of the amplifier are recorded on diagrams shown in Fig. 8 and 9. The oscillograms of the output voltages, which were measured between the load terminals and ground verified the switching speed. The barely visible voltage spikes are within the safe operating limit of the semiconductor components.

Actual modulation of the switching pulses occur in the magnetic bearing controller. The modulating circuitry contains a minor feedback loop and a pulse width modulator as shown in Fig. 9. The minor feedback loop maintains the bearing current or magnetic flux in proportion to the PID control signal by means of the pulse width modulator. Operation of the arrangement is apparent from the diagram in Fig. 10. The bearing current wave form shows the 40 kHz triangular switching ripple which is superimposed on a 500 Hz sine wave with a DC offset.

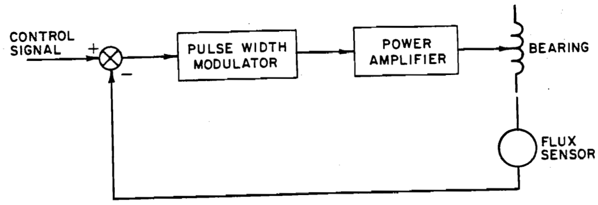

FIGURE 9

BLOCK DIAGRAM OF THE PULSE WIDTH MODULATING CIRCUIT

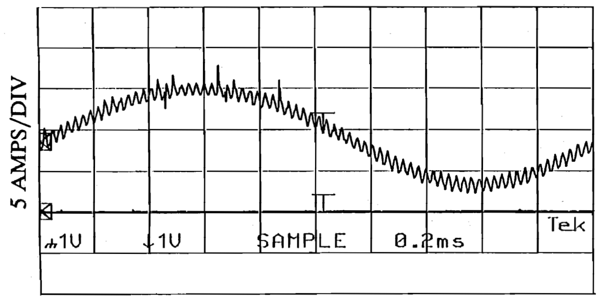

FIGURE 10

CURRENT IN A 0.4 mH LOAD DRIVEN BY POWER AMPLIFIER

References

[1] MASLEN, E., et al., "Pratical Limits to the Performance of Magnetic Bearings: Peak Force, Slew Rate, and Displacement Sensitivity, "Journal of Tribology, Vol.111, April 1989, pp.331-336

[2] MILLMAN, J., Micro-Electronics Digital and Analog Circuits and Systems, New York, McGraw Hill, 1979 ,pp.52-54

[3] MACOMBER, L., "Compute Thermal Resistance-The Key to Computer-Grade Capacitor Ripple," Electronic Design, Vol.12, June 1979, pp.80-83
[4] SEVERNS, R., "MOSFET Rise to New Levels of Power", Electronics, May 1980, pp.145-149
[5] LABERKRACK, J., Linear/Switch Mode Voltage Handbook, Motorola, 1982, pp.81

Nomenclature

Symbol		Meaning
i_b	=	current in bearing quadrant coil
g	=	air gap between bearing stator and rotor
f_{pwm}	=	frequency of pulse with modulator
F_b	=	force produced by bearing axis on rotor
F_{peak}	=	peak dynamic force
L_b	=	inductance of bearing
n	=	shaft speed (rpm)
N	=	number of turns in bearing quadrant
P_q	=	power loss in a bearing quadrant including amplifier losses
R_b	=	resistance of bearing quadrant coil
V_b	=	voltage applied across bearing quadrant coil
V_S	=	voltage of power supply
V_t	=	voltage across transistor

Appendix A

The rate of force change in a bearing can be determined in the following manner.The force in a bearing is given by the following equation as long as the reluctance of the core is small compared to the air gap.

$$F = \frac{A\mu_0(Ni_b)^2}{4g^2} \qquad \text{(A1)}$$

The partial derivative of bearing force with respect to current becomes;

$$\frac{\partial f}{\partial i_b} = \frac{A\mu_0 N^2 i_b}{2g^2} \qquad \text{(A2)}$$

The rate of current change in the bearing is determined by the voltage across the bearing coil and the inductance of the bearing.

$$\frac{di}{dt} = \frac{V_b}{L_b} \qquad \text{(A3)}$$

If the reluctance of the core material is again neglected, the inductance of the bearing is:

$$L_b = \frac{N^2\mu_0 A}{2g} \qquad \text{(A4)}$$

Combining equation (A3) with (A4) and mulptiplying by equation (A2) yields the bearing force rate change:

$$\frac{dF}{dt} = \frac{\partial F}{\partial i_b}\frac{di_b}{dt} = \frac{i_b V_b}{g}$$