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RF Note 112
A. McGilvra

February 1992

Hall Effect Current Transducer Note
Hall Effect Current Transducer Schematic

Appendix A: Technical Bulletins
1. Using an F.W. Bell Hall Generator
2. Temperature Compensating Hall Generators

Appendix B: Survey of DC Current Measurement Techniques for High Current Precision Power Supplies

HALL EFFECT CURRENT TRANSDUCER NOTE

2/21/92 A. McGilvra

We use non-intercepting current transducers to measure current in the 20 Volt/20 Amp and 10 Volt/100 Amp beamline power supplies. The transducers are based on a Hall effect sensor and coil in a feedback loop which makes a zero-flux transducer.

Several problems were encountered with the transducers that we bought from Telcon. The offset adjustment pot was single turn and hard to adjust. The unit drifted with time and would have to be readjusted after power cycling.

We designed a circuit using the Hall sensor and coil from a Telcon unit. We used a better op-amp (OP177EP), a ten-turn pot, and a modified nulling and excititation circuit.

The nulling circuit should be directly across the sensor as shown in the schematic. The excitation should be a current source. These were described in Technical Bulletin #H-100 published by F.W. Bell, a copy is attached. These two changes made the circuit much more stable with time.

Adding bypass caps across the op-amp reduced the problem of hysteresis and offset changes after power cycling.

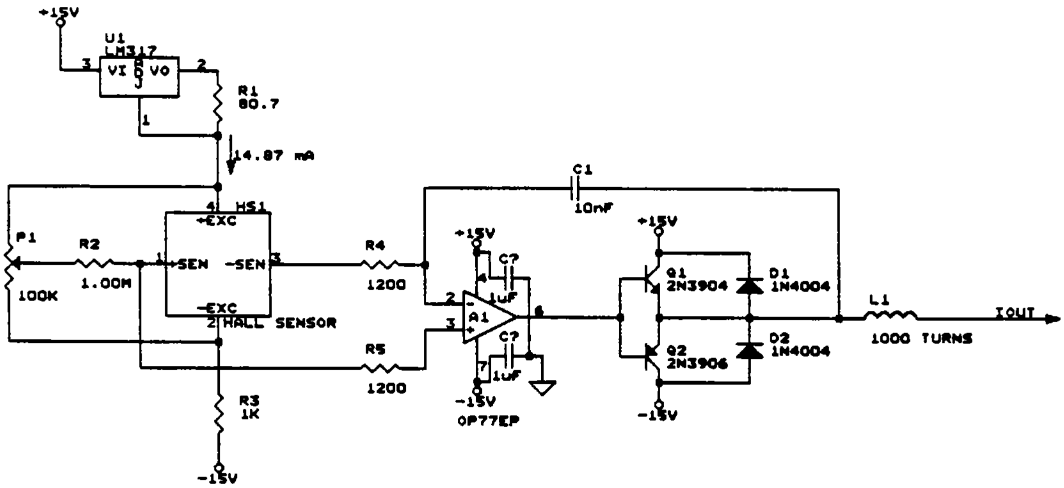
The circuit was sensitive to the external field produced by the K1200 main magnet. We measured 4 Gauss near the circuit with the magnet at 470, 700. The resultant offset was 26 mA of primary current, or 26 uA of transducer current. We put the circuit in a mu-metal box and then there was no noticeable effect until the field got to around 30 Gauss.

The circuit is still too sensitive to temperature. This manifests itself as a change in the offset when the temperature changes. When the temperature was raised by 10 °F the offset changed by more than 50 mA. This means that if the temperature changed by only one degree the 20V/20A PS would be out of its 1-in-10^4 specification (5 mA). There are ways to temperature compensate a Hall effect transducer as described in the F.W. Bell Technical Bulletin #H-102, a copy is attached.

The circuit is also sensitive to sudden changes in the primary current. This causes a non-zero flux to develop in the core which leaves a residual field that the sensor detects. The result is an offset error. When the units are used with a superconducting magnet load this is not a problem because the current can not change suddenly. However, when the units are being repaired or calibrated on the bench with a simple resistive load then special care has to be taken to never suddenly change the primary current. If there is a sudden change then the offset must be readjusted.

APPENDIX A

HALL EFFECT CURRENT TRANSDUCER



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Date: February 21, 1992 Sheet of

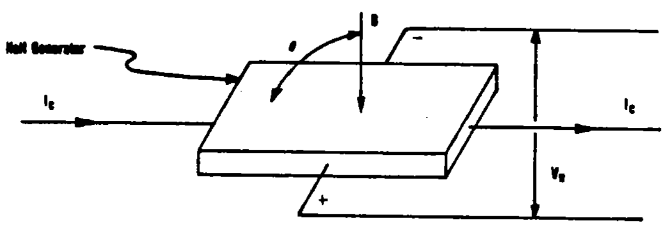
F.W. BELL — TECHNICAL BULLETIN — No. H-100

Subject: USING AN F.W. BELL HALL GENERATOR

CAUTION! To avoid possible permanent damage to the Hall generator, please read the following instructions carefully before making connections to a power supply.

INTRODUCTION

A Hall generator is a four-terminal, solid-state device capable of producing an output voltage, V_H, proportional to the product of the input current, I_C, the magnetic flux density, B, and the sine of the angle between B and the plane of the Hall generator.



$$V_H = K_{HOC}\, I_C\, B \sin\theta \text{ or if } \sin\theta = 1 \text{ (i.e., } \theta = 90°)$$
$$V_H = K_{HOC}\, I_C\, B \text{ or } V_H = \gamma_B\, B$$

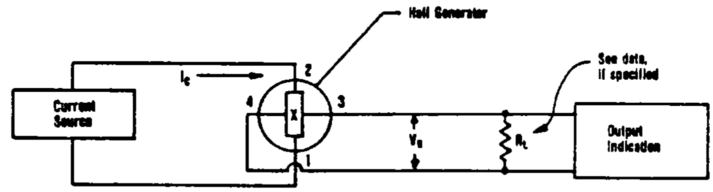
where: V_H = Hall output voltage, mV
$K_{HOC} = \gamma_{IB}$ (open circuit product sensitivity constant), mV/mA kG
γ_B = magnetic sensitivity (loaded or unloaded) at a specified control current, mV/kG
I_C = control current, mA (ac or dc)
B = magnetic flux density, kG (ac or dc)

Figure 1

A reversal in the direction of either the magnetic field or the control current will result in a polarity change of V_H. A reversal in the direction of both will keep the polarity the same. By holding the control current constant, the Hall voltage may be used to measure magnetic flux density. Multiplication may be accomplished by varying both the control current and the magnetic field.

CONNECTING THE HALL GENERATOR

The following schematic diagram illustrates the proper connections for the Hall generator:



Lead connections: 1 and 2 are control current (I_C) leads
3 and 4 are Hall voltage (V_H) leads

Polarity: AWG 34 red ($+I_C$), black ($-I_C$)
blue ($+V_H$), yellow ($-V_H$)

AWG 36 neutral ($+I_C$), green ($-I_C$)
red ($+V_H$), neutral ($-V_H$)

Figure 2 Hall Generator Circuit Configuration

Refer to the Hall generator specification sheet for color code or polarity. If a loading resistor, R_L, is specified, then it must be added to the output circuit as shown above to obtain the specified output.

Current Source

A constant current supply is recommended for applications requiring fixed control current. This eliminates effects of input resistance changes resulting from temperature or field variations (magnetoresistance effect). A "brute-force" constant current source may be made by connecting a large resistor (30 times R_{in} or higher) in series with a battery or constant voltage power supply. In any case, the short-circuit current should be within the maximum current rating of the Hall generator. The control current may be either ac or dc. This is determined by the nature of the field and the type of output signal desired. Refer to specification sheet for proper current for device being used.

Output Indicator

The Hall output voltage, V_H, may be observed on any suitable instrument such as a millivolt meter, oscilloscope, or recorder. The input impedance of the instrument should be greater than approximately 1000 ohms.

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CAUTION! ISOLATION REQUIRED!

Since the four Hall generator leads connect to four points on a semiconductor plate having different potentials, no two leads can be connected together without upsetting the operation. Therefore, the current source and the output indicator cannot have a common connection, but must be isolated from each other. One or the other, but not both, may be grounded.

Misalignment (Null) Voltage Compensation

In the manufacturing of the Hall generator, the Hall voltage contacts are placed on the semiconductor plate as accurately as possible so that very little output voltage will exist when there is no magnetic field present. For many applications, this resistive null voltage is low enough to be neglected, but for low field applications, it may be appreciable compared to the Hall output voltage, V_H. If this is the case, a null voltage balancing network such as that in Figure 3 will make it possible to reduce the resistive null voltage to zero. The fine control may not be required.

Affects of Residual Magnetism

Care should be taken to ensure that what appears to be an offset voltage of the Hall generator is not really the result of a residual magnetic field. Any magnetic material with a residual field in close proximity to the Hall generator could effect a slight Hall output voltage, V_H. Items such as fixtures, jigs, probes, metal tables, metal cabinets, etc., are potential sources of residual magnetic fields. Even the earth's magnetic field (approximately ½ gauss) could cause an undesirable "offset" voltage. The circuit in Figure 3 can be used to zero out many of these voltages.

SPECIAL APPLICATIONS

Additional information on specific applications of Hall generators may be obtained by writing or calling F.W. Bell, Inc.

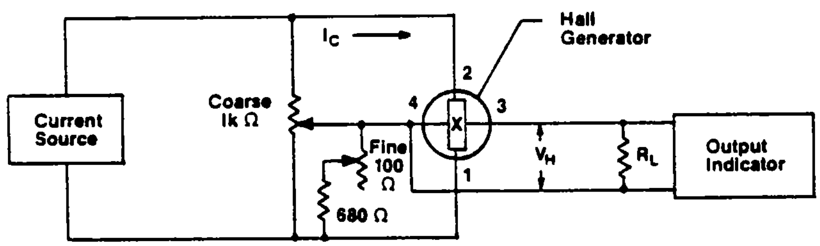

Figure 3 Misalignment (Null) Voltage Compensating Network

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Subject: Temperature Compensating Hall Generators

By J. A. Dierker
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Temperature effects can rarely be ignored in the application of a Hall generator. Selecting a generator with a low-temperature coefficient of sensitivity may suffice, but compensation is often required. If the designer understands the parameters involved, their effects on the Hall device output, and knows the ambient operating conditions, applying the compensating networks is simple. Due to the wide range of parameter variations among Hall generators of the same type, however, it is difficult to have one compensation network work for all generators that are used.

The compensation circuits offered below represent the basics; the total possibilities are limited only by the user's imagination. Only N-type indium arsenide Hall generators will be considered. Conduction is by majority carriers, no semiconductor junctions exist, and all contacts to the semiconductor material are ohmic.

An indium arsenide Hall generator usually exhibits a positive temperature coefficient of resistance (α_T) over its normal operating range. If a constant voltage supply provides the generator control current, this current decreases as the Hall plate temperature increases. The result is an output sensitivity inversely related to temperature. A constant current supply eliminates the effects of Hall generator input resistance changes. This condition can also be approximated by a voltage supply with high output impedance. If conditions require a low-impedance voltage supply, the designer who wants to compensate for α_T must be aware of a sign reversal on it, as shown in Figure 1.

For most Hall generators with product sensitivities below 0.2 volts per ampere kilogauss, the sign change on α_T takes place at temperatures above the normal operating range. In this case, the compensating networks discussed in the following sections may be used. Higher sensitivity devices exhibit the sign reversal within their operation range, however (see Figure 1). This behavior can make compensation very difficult, and the designer may have to accept the input resistance temperature effect or change to a higher source impedance.

Sensitivity variation

The relation of electron mobility to temperature causes Hall generator sensitivity to decrease as temperature increases. This phenomenon is independent of the temperature coefficient of resistance effect. Normally, Hall generators with higher sensitivities exhibit higher temperature coefficients of sensitivity (β_T) at constant current. The output variation is approximately linear from −60° to 60°C. Above 60°C, the rate of change increases rapidly (see Figure 2).

Several methods exist for compensating for temperature variation of the parameters discussed above. Since all the suggested procedures require trial-and-error calculations and testing, only the overall configuration is described.

A positive temperature coefficient (PTC) resistor can be used as a compensating element (Figure 3). Nickel or copper wire resistance are possibilities, or PTC silicon resistors may be considered. Nickel exhibits a positive temperature coefficient of about 0.5%/°C; copper, 0.4%/°C; and silicon resistors, about 0.7%/°C.

Negative temperature coefficient (NTC) thermistors are the most widely used devices for compensating Hall generator temperature drift (Figure 4). This configuration with a voltage supply causes the control current, I_C to increase as the temperature rises.

Another compensation method, using an NTC thermistor, is shown in Figure 5. In implementing this concept, the designer must make sure that two conditions exist. Ample output must be maintained across resistor R_1, and the total load, R_1, presented to the Hall generator must not adversely affect output linearity. (Consult the Hall generator manufacturer if linearity is a critical parameter.) If amplification is required, an alternative to the circuit in Figure 5 would be to place a thermistor circuit in the feedback network of the amplifier.

Single thermistor or PTC resistor circuits usually exhibit good high-temperature correction, or vice versa. Improved compensation over a wider temperature range may be achieved through use of two temperature sensor networks. The two may be a combination of those mentioned above.

The following example (Figure 6) illustrates the results of negative temperature coefficient thermistor compensation in the Hall generator input circuit. The Hall generator has the characteristics listed below:

Input Resistance (R_{in}), 4Ω
Magnetic Sensitivity (γ_B), 30 mV/kG,
(At I_C = 100 mA and T = 25°C)
Temperature Coefficient of Input
Resistance (α_T), +0.2%/°C
Temperature Coefficient of Magnetic
Sensitivity (β_T), −0.22%/°C

Improved Hall generator temperature drift is desired between 25° and 60°C. Uncompensated, the generator sensitivity at 60°C would decrease approximately 8% from its value at 25°C. A first attempt to improve this temperature drift might be the circuit shown in Figure 6 (b). The real problem is to select proper values for R_1 and R_T.

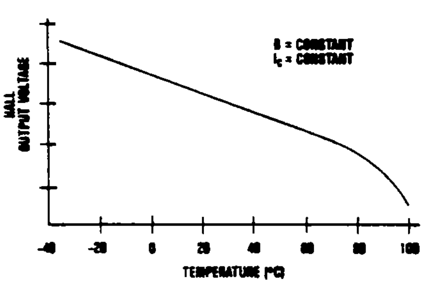

Figure 1. "Worst case" temperature characteristics of input resistance.

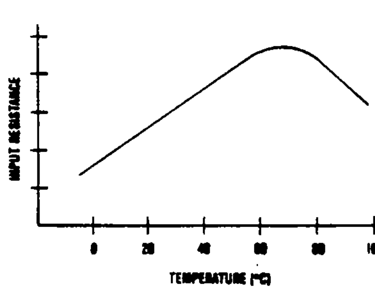

Figure 2. The sensitivity of a Hall generator decreases with temperature.

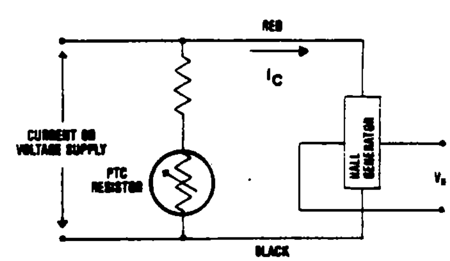

Figure 3. A positive temperature coefficient thermistor can be used in temperature compensation of Hall generators.

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Increasing the Hall generator control current by about 8% at 60°C will eliminate the drop in sensitivity due to β_T. The next step is to calculate a thermister-resistor parallel resistance, R_{TOT} which will supply 108 mA control current at 60°C. (Changes in resistor R_1 and in the Hall generator input resistance are small and will be ignored throughout this example).

$$R_{TOT} = (V_S/I_C) - R_{in}$$
$$= 5/(0.108) - 4 \cong 42.3\Omega$$

Since thermistor resistance will be selected at 60°C, R_T will generally be high in value at 25°C. Assuming that R_T will be moderately high, let R_1 be 47Ω. If our assumptions are correct, the Hall generator control current will be near 100 mA at 25°C. Next, the thermistor resistance at 60°C must be calculated:

$$R_T = \frac{R_1 (R_{TOT})}{R_1 - R_{TOT}}$$

The table below gives a comparison of sensitivity change for both the uncompensated and compensated circuits (a thermistor temperature coefficient of −3.9%/°C was used). A higher R_1 might be selected for better low-temperature response.

T (°C)	γ_B (mV/kG)	R_T (Ω)	R_{TOT} (Ω)	I_C (mA)	γ_B (Comp.) (mV/kG)
−10	32.3	6447	46.66	98.7	31.88
+10	31.0	2636	46.18	99.6	30.88
+25	30.0	1440	45.51	101.0	30.30
+40	29.0	827	44.47	103.2	29.93
+60	27.7	423	42.30	108.0	29.92

Test results on sensitivity with compensated and uncompensated circuits.

Zero field residual voltage

In most cases, the zero field residual voltage (V_{MT}) temperature drift is separate, but not distinguishable, from sensitivity drift. When magnetic flux densities are high, the effect of V_{MT} temperature drift may be negligible. In situations of low magnetic field, the temperature drift of the residual voltage may be equal to or larger than the sensitivity change. The inability to consistently and accurately predict the rate and direction of drift in this offset voltage is the most difficult part of compensation.

Since the major portion of the V_{MT} is caused by unbalanced output contact positioning (misalignment voltage), the characteristics of the temperature drift may be predicted with fair accuracy. For devices with a room temperature V_{MT} greater than 500 μV, the magnitude of the zero field residual voltage increases at a rate approximately equal to the temperature coefficient of resistance, α_T. Units with V_{MT} between 100 and 500 μV normally exhibit an increase in residual voltage magnitude, but at rates varying within ±100% of the resistive coefficient. Below 100 μV, the task of predicting the offset drift becomes difficult because the direction and rate are random. A circuit that can be used to compensate for the zero field residual voltage temperature drift is shown in Figure 7.

The thermistor shown compensates for a positive-going temperature drift (with respect to the output terminals). If the zero field residual voltage varies negatively with increasing temperature, the thermistor network and R_1 must be interchanged.

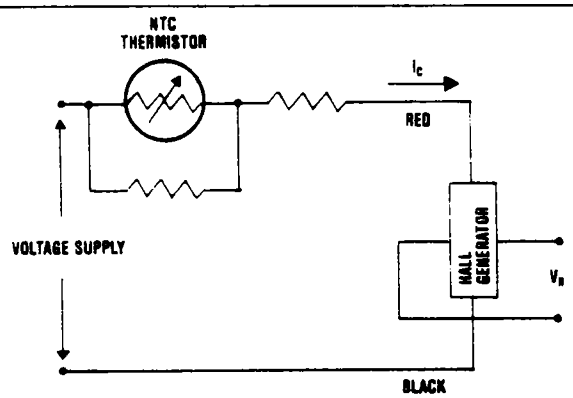

Figure 4. Hall generator can be compensated at its input by a negative temperature coefficient thermistor.

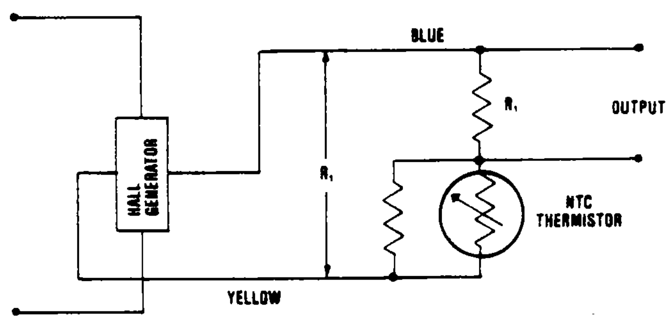

Figure 5. Temperature compensation of a Hall generator using a negative temperature coefficient thermistor.

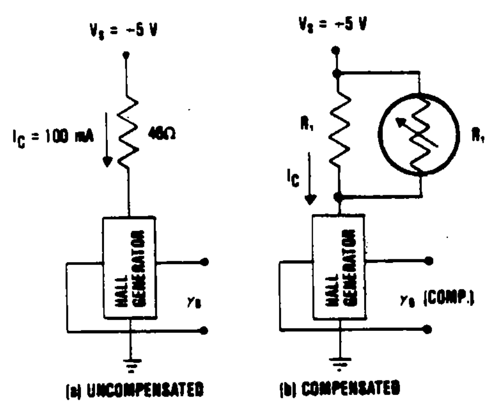

Figure 6. The use of a negative temperature coefficient thermistor as a compensator in the Hall generator input.

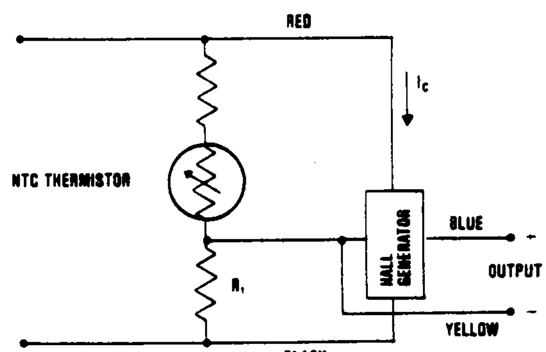

Figure 7. Compensation of zero field residual voltage temperature drift of Hall generators.

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MIL-STD-793-1 (WP) "Definitions Letter Symbols, Color Code and Circuit Symbol for Devices, Hall Effect"

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APPENDIX B

Survey of DC Current Measurement Techniques for High Current Precision Power Supplies

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SURVEY OF DC CURRENT MEASUREMENT TECHNIQUES FOR HIGH CURRENT PRECISION POWER SUPPLIES

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Abstract

This paper will investigate the various current measurement techniques typically found in precision DC current regulated power supplies used in particle accelerator applications. The relative strengths of each type will be indicated.

Introduction

Direct current power supplies used in particle beam accelerator applications are primarily operated in the constant current mode. The reason for this is simple enough. The direction and velocity of charged particles are affected by DC magnetic fields. This property allows the bending and focussing of particle beams, and is a fundamental principal of particle accelerator design and operation.

DC magnetic fields are created by passing direct current through precisely designed and manufactured magnets. The magnetic field will vary as the current flowing through the magnet varies. As it is imperative that the particle beam be stable for effective accelerator operation, it is therefore imperative that the DC power supply provide stable and precisely regulated current.

The key to precise current control is the ability to precisely measure the current. The degree of current precision required is dependent on the accelerator physics of the particular application. For instance, the current precision specification for an achromatic dog leg in a beam line may be considerably less than that for the dipole and quadrupole magnets found in an antiproton accumulator where resonances of up to order eleven are avoided. Accuracy is not important in most particle accelerator applications. Repeatability and long-term stability are the parameters which matter.

Many methods of current regulator design are possible. The most common method controls the power supply voltage by integrating the difference between a current reference signal and a current feedback signal. In order to preserve signal integrity these signals need to be in the range of 1 to 10 volts. This paper is concerned with the precision of the current feedback signal.

Only DC characteristics will be discussed. Specifications for current measurement devices often include misused and nonstandard terms. Appendix A details some definitions which will help with the clarity of this paper.

Measurements, Equipment, and Techniques

Many current measurement devices were measured over the last few months. This included measuring the effects of temperature, time, leakage flux, humidity, mechanical stress, and external noise on the current measurement precision. In addition, accuracy, frequency response, noise levels and linearity were measured for most devices. The measured devices included shunts, Hall effect current sensors, current transductors, and direct current current transformers (DCCTs). Measurements were made at currents from 10 amps to 8,000 amps.

All measurements were made using standards consistent with the precision of the device being measured. Precision was monitored by measuring a high precision reference. For instance, when measuring the temperature coefficient (T.C.) of transfer impedance (Zt) of a 10 amp shunt, the current was monitored using a 2.5 ppm/°C shunt in a constant temperature chamber. The voltages were measured using a Prema 8½ digit volt meter in a constant temperature chamber.

The equipment used for these measurements included two home-made temperature chambers. The first one held the temperature of measurement equipment constant at 25±1 °C. The second was controlled by a personal computer to temperature cycle devices under test. Most voltage and temperature measurements were made using a Prema 6047 precision volt meter connected to a precision signal scanner. Various Dynapower built precision current sources were used to stimulate the devices under test. All temperature probes were 100 ohm platinum resistors.

Temperature cycle tests were performed on all the devices presented here. The computerized test system was capable of continuously scanning up to 20 various types of measurements, while controlling the temperature cycle. A typical temperature cycle is shown in Figure 1 .

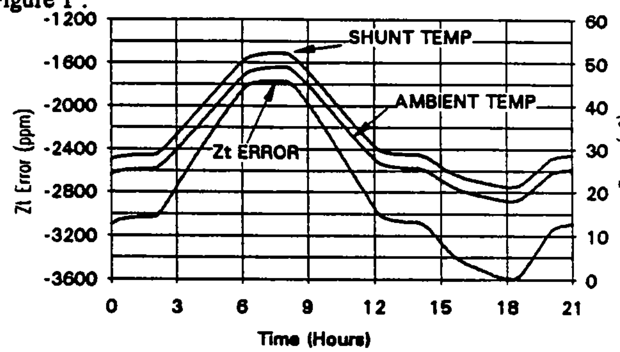

Figure 1: Transfer impedance error and ambient temperature of a 10 amp shunt measured at 10 amps versus time.

This continuous measurement technique provides a lot of information about the device under test. It not only gives a highly detailed temperature coefficient curve, it also makes the resolution of the measurement immediately obvious. If the temperature is cycled up and down, the resulting temperature retrace curve gives a good indication of the repeatability and long term stability of the device. Figure 2 shows a temperature retrace curve for a wire wound resistor.

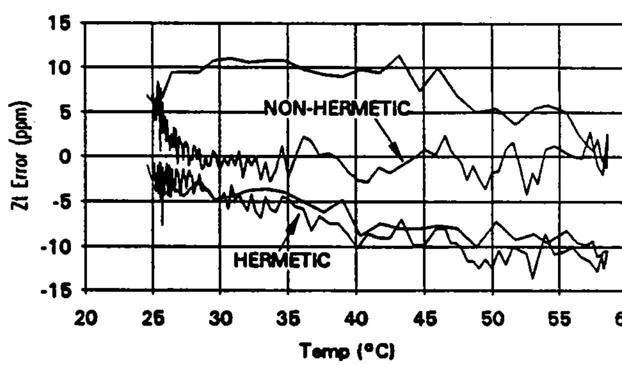

Figure 2: Transfer impedance error of two 5,000 ohm, 1 ppm/°C wire wound resistors versus ambient temperature.

The large open curve is explained by the manufacturer as a humidity effect. The theory is that the enamel insulation absorbs water and stresses the wire causing the resistance to shift. The open loop results because the water absorption lags the temperature. This resistor would not make a very repeatable resistor in a constant temperature arrangement even though its temperature coefficient is better than 1 ppm/°C. The same type of open loop curve may also result from placing temperature probes in a location which leads or lags the temperature sensitive part of the device. For this reason, temperature must be run up and down slowly when looking for temperature retraceability. Figure 2 also shows a temperature curve of the same type of resistor in a hermetic package. Note that the up and down curves can not be distinguished.

The temperature retrace curves can also be used to compare the long term stability of devices. If a device does not return to the same value after mild temperature cycles it probably will not have good long term stability. Figure 3 shows a temperature retrace curve for a prototype 10 amp 0.5 volt metal foil shunt. This device will obviously not have long term stability on the order of 1 ppm/month.

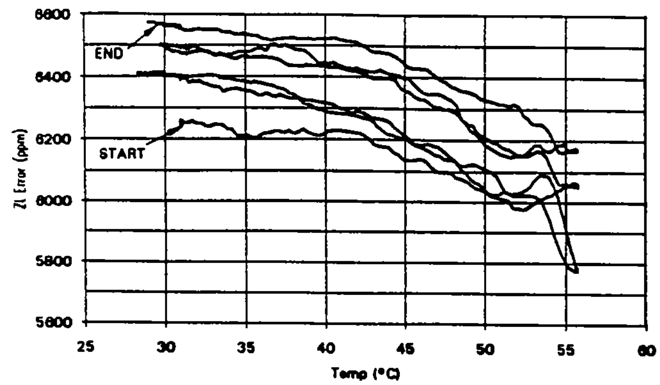

Figure 3: Temperature retrace curve for a prototype metal foil shunt, measured at 10 amps.

A potential problem when temperature cycling low output voltage devices is thermal EMF in the test probes. This can become a problem if dissimilar metals are used in the connection circuitry or if there are temperature gradients in the connections. If the precision of the measurement needs to be better than about 1 uV, care should be taken in making test connections.

Characteristics of Current Measurement Devices

This section will discuss the characteristics of current measurement devices used in precision high current power supplies. Three basic categories of devices will be discussed here: devices which use Ohms law (V=I*R), called current shunts, devices which use Amperes law (H*dl = I), called magnetic current sensors, and devices which use both, called DCCTs.

Current shunts

Shunts are differentiated from resistors by the fact that they are exclusively designed to measure current. The major factors which affect the precision of a shunt are: the temperature coefficient of transfer impedance, power dissipation, low output signal, and in many applications, the need for isolation from the current being measured.

The first three of these factors are interrelated. Low voltage outputs (typically 50 mV) are chosen in order to keep power dissipation low in a high current shunt. Low power dissipation minimizes the effect temperature coefficient of Zt has on linearity. In order to measure a 50 mV output to 10 ppm, an amplifier with 500 nV precision is required. This is not practical. At lower currents, a higher output voltage may be selected. This allows the output to be measured more precisely. At currents below 100 amps, shunts are used as primary standards by many calibration laboratories.

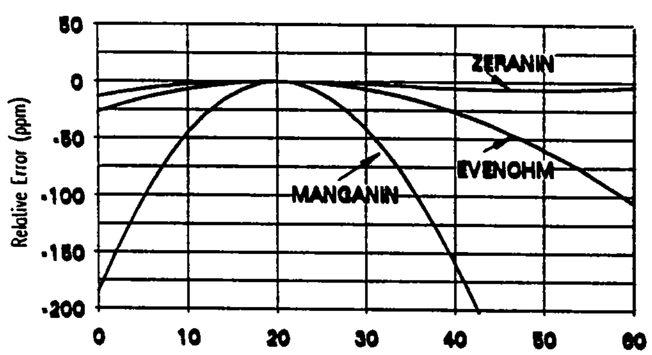

Figure 4: Resistance change of Zeranin, Manganin and Evenohm resistance alloys versus temperature. These curves have been normalized to have zero T.C. at 20 °C. The unnormalized curves will be shifted from these by as much as 50 °C.

Consider a 2,000 amp 50 mV shunt with a thermal impedance of 0.25 °C /watt. It has a constant temperature coefficient of -10 ppm/ °C and will need to operate in an ambient of 25 to 55 °C. This shunt will shift 300 ppm over the ambient temperature range. Its temperature will rise 25 °C above ambient at full current. This temperature rise will cause it to shift -250 ppm in the few minutes it takes to warm up. The linearity of this device could be specified from ± 32 to ± 125 ppm, depending on the definition of linearity used.

A real shunt would have an additional complication, the temperature coefficient is not constant. The shape of a Zt error versus temperature curve is characteristic of the resistance alloy being used. There are many manufacturers making shunts using the above alloys or variations of them, as shown in Figure 4. Manganin shunts are usually specified at 0.25 % accuracy with a T.C. of ± 15 to 20 ppm/°C. Evenohm and Zeranin shunts are used for higher precisions and lower current applications and have tolerances from 100 to 1,000 ppm and T.C. specs of ± 1 to 5 ppm/°C. The following is a list of shunts that were measured.

	MANUFACTURER'S SPECS				MEASUREMENTS		
Resitance Alloy	Current Rating	Voltage Rating	Accuracy	T.C. of Zt (ppm/°C)	True Value	T.C. of Zt (ppm/°C)	Current Used
Manganin	12.5A	50mV	0.25%	+/-20	2.330%	108	10A
Manganin	12.5A	50mV	0.25%	+/-20	-0.713%	40	10A
Manganin	10A	50mV	0.25%	+/-20	-0.240%	68	10A
Manganin	10A	50mV	0.25%	+/-20	-0.250%	59	10A
Even ohm	50A	500mV	0.02%	+/-2.5	0.015%	1.5	10A
Metal Foil	10A	0.5V	NONE	NONE	0.560%	12	10A
Zeranin *	100A	1V	0.10%	+/- 5	0.225%	13.7	100A
Zeranin *	133.3A	1V	0.10%	+/- 5	-0.495%	11.8	100A
Zeranin *	100A	1V	NONE	+/-3	4.462%	4.1	100A
Manganin	100A	50mV	0.25%	+/-20	0.119%	-4.65	100A
Manganin	100A	50mV	0.25%	+/-20	-0.085%	12	100A
Manganin	100A	50mV	0.25%	+/-20	-0.069%	16.3	100A
Manganin	8000A	50mV	0.25%	+/-20	-1.400%	15	4000A
Manganin	8000A	50mV	0.25%	+/-20	0.890%	35	4000A
Manganin	8000A	50mV	0.25%	+/-20	1.680%	165	4000A

*Table 1: Various shunts measured and their manufacturers specs. The T.C. of Zt was measured from 25 to 50 °C and the absolute error was measured at 30 °C. (* indicates water cooled)*

Note that only 6 of 13 devices met specifications for accuracy and 5 of 14 for T.C. of Zt. This means that shunts need to be measured by the purchaser to insure performance.

Magnetic Current Sensors

Magnetic current sensors comprise devices which measure the magnetic field generated by the current flow. These include: Hall effect current sensors, magnetic modulator devices, NMR current sensors, SQUID current sensors, magnetoresistive devices, and magnetostrictive devices.

Hall Effect Devices: When a magnetic field is applied to a conducting or semiconducting material in which a current is flowing, a voltage will be developed across the sides of the material. This is known as the Hall effect. A Hall effect current sensor consists of a magnetic core with one or more gaps in which Hall effect devices are situated. The current to be measured passes through the center of the core creating magnetic fields in the gaps. If a constant current is supplied to the Hall effect device, then a voltage is produced that is proportional to the magnetic field in the core. There are devices available commercially which can be used to measure currents up to 4,000 amps.

The major factors affecting the precision of Hall effect current sensors are: temperature sensitivity, linearity, leakage flux sensitivity, and sensitivity to control current.

The temperature sensitivity has two components: temperature coefficient of transfer impedance and temperature coefficient of offset. The relation of electron mobility to temperature causes Hall generator sensitivity to decrease as temperature rises. Normally, Hall generators with higher sensitivities have higher temperature coefficients [2]. This fact causes Hall effect current sensors to have temperature coefficient of transfer impedances in the range of 25 to 1,000 ppm/°C which are not constant with temperature, as illustrated in Figure 5.

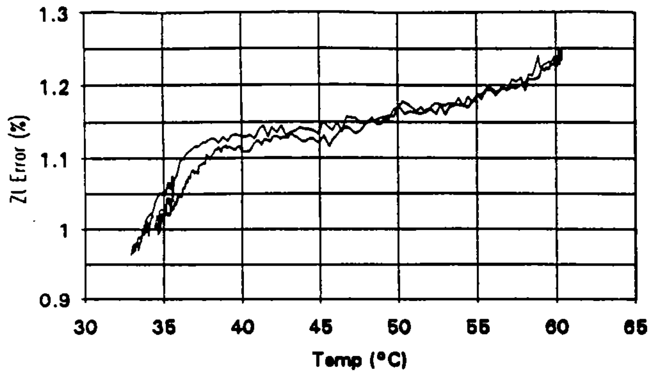

Figure 5: Temperature retrace curve of a 100 amp Hall effect current sensor, measured at 10 amps.

Hall sensors exhibit an offset at zero magnetic field mainly due to misalignment of the output terminals. This offset has a temperature coefficient. The size of the error caused by temperature coefficient of offset depends on the current being measured. At 400 amps and above the T.C. of offset could be as good as 10 ppm/°C. At lower currents it would get worse.

The temperature sensitivity of Hall effect devices can to some extent be compensated for with external electronic circuitry [2]. This can be complex since the temperature coefficients are nonlinear and vary from device to device.

Ampere's law implies that the line integral of the magnetic flux around a core will equal the total current passing through the core. Since the Hall effect device only measures the field at a point, its precision depends on a uniform field. Thus, leakage flux (see Appendix definition) irregularities introduce errors into the measurement. For this reason leakage flux should be minimized by centering bus current and keeping return buses far away. For low currents, leakage flux is not much of a problem, but at higher currents (above 1,000 amps) this must be carefully considered or errors over 1% could occur.

Sensitivity is directly proportional to the control current in the Hall generator. The control current must then be controlled to a better precision than the Hall generator.

Hall effect current monitors can make reliable and inexpensive instruments when used at the level of 0.5 to 5% precision. They can have response times on the order of 1 uS and can be utilized in clamp-on type portable measurement systems.

Second Harmonic Magnetic Modulators are members of the magnetic amplifier family. They are used in very high sensitivity magnetic amplifiers which have very high resolution and long term stability. Long term drift of better than 100 uA/(meter*month) have been obtained in particle accelerator beam current monitors which are now commercially available [7].

The low offset error characteristics of a second harmonic magnetic modulator are due to the near perfect symmetry of a hysteresis loop. This symmetry is unaffected by temperature and time and is the main reason for the low offset error character of the device.

Consider a single saturable reactor (SR) with an AC and a DC winding. Start with the DC winding open. On the AC winding is a sine wave voltage excitation causing the SR to go far into saturation. The current in the AC winding will be symmetrical due to the symmetry of the hysteresis curve. Thus, the frequency content of the current signal will have only odd harmonic components. When a DC current is applied, the current in the AC winding will become asymmetric. This now creates an even harmonic signal proportional to the DC current. Prior to DC saturation the relation between the DC current and even harmonic current is approximately linear. The second harmonic current is usually monitored, but any of the even harmonics could be used. The AC excitation does not have to be a sine wave, it just has to be free of even harmonics.

A single core second harmonic magnetic modulator produces AC voltage on the DC winding which can cause problems in the current source. A second problem is that the AC excitation must have very low second harmonic content. To reduce these problems, two identical SR cores are typically employed. They are connected such that the AC windings are in parallel, with equal turns. The DC winding is connected such that the signals from the AC windings cancel. The need for low, even harmonic content in the AC excitation voltage is lowered by the degree of matching of the SR cores. The current is then monitored for second harmonic content as before.

At low DC currents the second harmonic is so small relative to the odd harmonics that it is hard to detect. This problem is easily solved in the twin core magnetic modulator by the addition of a winding which is in series with both cores and in the same sense as the DC winding. If the cores are matched perfectly there is no odd harmonic voltage on this winding, only even harmonics.

This type of magnetic modulator has been used to resolve currents of less than 1 uA when using 3 inch diameter supermalloy tape wound cores. This same unit would only operate up to about 1 amp.

When used in this way, a second harmonic magnetic modulator is limited to measuring currents below a few amps. When used as a null detector in a well designed DC current ratio comparator, it has been used to measure current ratios to better than 1 ppm at 20,000 amps [6].

NMR, SQUID, magnetoresistive and magnetostrictive devices could be used to make current measurement devices, but have not been applied to a large degree to current measurement in magnet power supplies.

Direct Current Current Transformers

DCCTs use magnetic current sensors to extend the frequency range of an AC current transformer down to DC. By using a specific turns ratio, a smaller secondary current, which is galvanically isolated and approximately proportional to the primary current, is generated. This secondary current can then be accurately measured with a low power, high stability shunt to give a scaled readout of the primary current.

In a DCCT the magnetic measurement device is operated in a zero field environment. In this configuration only their zero offset characteristics introduce errors in the output. Turns ratios in DCCTs range from 500 to 10,000 : 1. The secondary currents range from 0.1 to 10 amps. In this range of currents it is possible to find shunts with precision on the order of 1 ppm.

The current measurement systems which will be discussed here are current transductors, Hall effect DCCT's, and second harmonic magnetic modulator DCCT's. Current transductors have been included here because they have most of the desirable characteristics of DCCTs.

Current Transductors are fundamental frequency magnetic modulators and members of the magnetic amplifier family. Their basic DC current sensor is a twin core saturable reactor, in which, at zero DC current, the total flux swing approaches the available maximum. Transductor designs vary widely, from the simple to the very complex. For a description of the basic series connected current transductor see Geyger [3].

The parameters which have the largest effect on the precision of a current transductor are line voltage, leakage flux, and temperature. The reasons that these parameters affect the precision are complex. The problems arise because of the resistance of the secondary winding and the fact that there are no core materials which make ideal saturable reactor switches. Even if there were perfect materials to use, a current transductor cannot tell the direction of the current. Also, a transductor cannot measure current well near zero, unless offset compensation is added [5].

Most current transductors use the AC line voltage directly as a modulation source. In a simple series connected current transductor a 1% change in line voltage produces a change on the order of 0.05% in the output. Variations in line voltage will have a larger effect on the current ratio at lower currents. This can be improved with more complex designs.

Leakage flux affects the transductor as follows: as the secondary current makes a transition into ampere turn balance in one of the saturable reactors, the leakage flux causes a dispersion in the time at which the core comes out of saturation. This soft transition causes the ampere turn balance point to be different then if there were no leakage flux. This effect varies with current and thus affects the linearity of the transductor. Nonlinearities can be calibrated out by making measurements at the important currents. However, if the leakage flux environment is changed, i.e. the transductor or bus work moved, the calibration must be repeated.

Temperature affects the core's magnetic characteristics, causing errors by changing the point of ampere turn balance. The best current transductors use high performance shunts and regulated line voltage. The overall precision of transductors varies widely from the order of 0.01 to 2%. The reliability of a simple transductor is very good. Most applications of the current transductor are in the measurement of currents over 5,000 amps, where shunts start to become impractical.

Hall effect DCCTs use one or more Hall effect devices as null detectors. This current measurement system's precision is limited by the following: the temperature coefficient of offset of the Hall sensor, hysteresis of the magnetic cores, and leakage flux errors.

Without temperature compensation, the temperature coefficient of offset error of this type of DCCT can be as good as about 10 ppm of full scale/°C at currents above about 400 amps. This is calculated using a Hall device with a temperature coefficient of offset of 0.05 Gauss/°C, operated at 5 kG full scale. The current of 400 amps is the current needed to make a 5 kG field in a core with a 1 millimeter gap.

Hysteresis errors are illustrated as follows: using a 4 inch diameter supermalloy core with a coercive force of 0.003 Oersteds in the above application, the magnetic field in the gap with zero current could be as high as ±1 Gauss. This would correspond to an offset error of ±200 ppm, depending on the history of the field in the core.

To understand leakage flux effects, consider a 400 amp return bus 6 inches from the center of the core and in the same plane as the gap. If the core were 1/2 inch thick, this would cause a field in the gap of about 20 Gauss. This field corresponds to an error of 0.4% of full scale. This problem gets worse as the current rating gets higher.

Fortunately there are some fixes for this problem. The simplest is to add more Hall effect devices, placed symmetrically around the core. The voltages of these devices can then be summed giving a better approximation to Ampere's law. This can significantly reduce the sensitivity to leakage flux. By looking at the signals from each of the Hall effect devices, the sensing head can be better centered in the magnetic field, thus reducing the leakage flux. This is not without trade-offs. The added gaps lower the magnetic field, so more primary current is needed to get the same temperature coefficient of offset performance. At currents above 5,000 amps the leakage flux makes appreciable fields in the cores, shifting of the fields in the gaps resulting in nonlinearity.

To help with this problem the idea of multiple Hall effect devices is carried one step further. In commercially available Hall effect DCCTs, separate current feedback windings are used for each section of the core which contains a Hall sensor. The current in each part of the core is independent so it can null the field at each sensor. The currents are then summed into a common shunt to measure total secondary current. This type of Hall effect DCCT is commercially available at currents up to 350,000 amps.

Measurements were made of a 6,000 amp 4 channel unit of this type. The offset temperature coefficient of the head was -5.8 ppm of F.S./°C from 25 to 45 °C. The offset of the head did not retrace back to the same point after the temperature cycle. It changed by -0.48 amp at 25 °C. This corresponds to -80 ppm of FS. This could be an indication of a repeatability or long term stability problem. The gain temperature coefficient of the head was -14.6 ppm/°C from 27 to 36 °C.

The leakage flux sensitivity was measured by placing the head 12 inches from a 6,000 amp bus, with the bus parallel to the center line of the head. This caused the individual channels to shift 13%, -8%, -3.8% and -1.2% of full scale. These errors canceled very well causing a total offset shift of only -83 ppm of full scale at the output.

If this device is not moved, the leakage flux errors and nonlinearity errors can be calibrated out. The resulting performance at full scale would be on the order of 0.1% over time and temperature. This precision can be quite acceptable at high currents. In fact, if isolation is needed a Hall effect DCCT is as good as a manganin shunt with a very good isolation amplifier.

Second harmonic DCCTs consist of a second harmonic magnetic modulator and a current transformer inside a common secondary winding (see Figure 6). The magnetic modulator signal provides the DC balance and the current transformer provides the transient response.

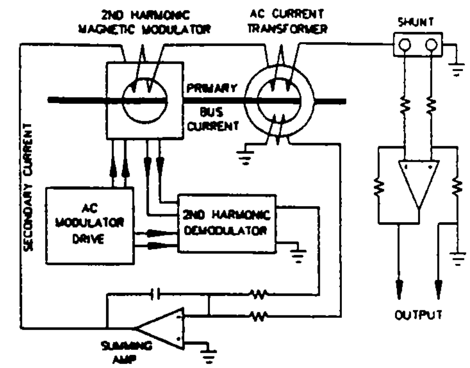

Figure 6: Simplified schematic of a second harmonic DCCT.

The precision of a second harmonic magnetic modulator is affected by: the long term stability of the modulator, the temperature coefficient of offset of the modulator, leakage flux, demodulator offset errors, modulator core memory errors, and secondary current measurement errors.

The long term stability of a second harmonic magnetic modulator is very good. Take for example a 10 inch core second harmonic modulator. The long term stability using a supermalloy core pair can be less than 50 uA turns per month. This 10 inch core could be used to make a 25,000 amp 2nd harmonic DCCT. In this case the 50 uA corresponds to 0.002 ppm of full scale per month.

The temperature coefficient of offset of a second harmonic magnetic modulator designed for a 4,000 amp DCCT has been measured to be zero with a resolution of better than 0.01 ppm/°C.

Without magnetic shielding leakage, flux will exist in the second harmonic modulator cores. This flux varies around the core. A line integral around the core should be zero if the DCCT is at amp turn balance. In practice, the variations around the cores do not cancel, even in very carefully wound modulators. Leakage flux does not start to cause significant errors (over 1 ppm) until the maximum leakage flux density in the core reaches the order of hundreds of Gauss. At this point, the errors start to increase rapidly due in part to problems with driving the lower impedance of the modulator. To demonstrate this, a 6 inch modulator (size used in an 8,000 amp DCCT) was placed 12 inches from a DC current bus with its axis parallel to the current. The current in the bus was ramped while the second harmonic was monitored (Figure 7).

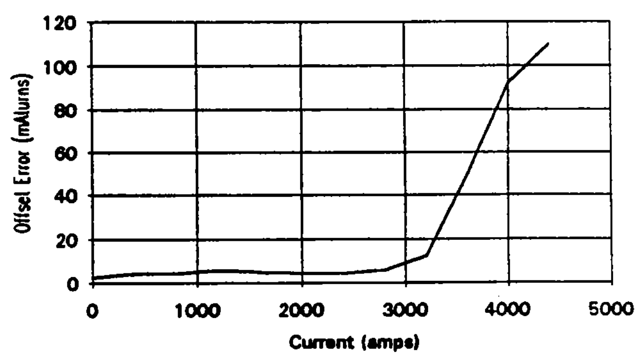

Figure 7: Offset error in an unshielded 6" second harmonic magnetic modulator versus current in adjacent bus used to induce leakage flux.

By putting a magnetic shield around the modulator cores, the cores can be shielded from the leakage flux [4]. To keep the shield from saturating, it must be under the secondary winding. As demonstrated by Figure 7, the modulator does not have to be shielded completely, the maximum magnetic field needs only to be below about 100 Gauss in the modulator cores. A shielded modulator identical to the one used to create figure 7 was tested in the same configuration at 8,000 amps. The output offset change was zero, to a resolution of 1.6 mA turns or 0.2 ppm of 8,000 amps.

The second harmonic demodulator can cause small amounts of error. Errors at the level of 0.05 ppm of F.S./°C have been measured. It is not believed that this is a fundamental limit.

Core memory effects may be a factor at low currents. Measurements have shown they can be less than 1 ppm in units sized for 1,000 amps.

A well designed second harmonic magnetic modulator DCCT makes a near perfect direct current current transformer. It is limited only by the shunt and amplifier that are used to measure the secondary current. Secondary currents range from 0.1 to 10 amps. In this range of currents a shunt and amplifier can have the precision of 10's of ppm over a 20 °C temperature range. By temperature controlling the shunt and amplifier the precision can be improved to the range of a few ppm over the same temperature range.

The very high precision of the second harmonic magnetic modulator is demonstrated in the use of this technology by national standards laboratories. Based on this technology, the National Research Council of Canada (NRC) has developed a DC current ratio standard for use at 20,000 amps [6].

Comparison

The choice of a precision current measurement device depends on many factors. In order to point out some of the factors involved, consider a 3,000 amp dipole bus power supply. Two currents are important in this application, 3,000 amps and 300 amps. The power supply will perform only slow ramps, less then 10 amps/second. The line voltage will vary ± 10% from nominal. The ambient temperature will be 20 to 45 °C. A single device is to be used to measure the current at both 3,000 amps and at 300 amps. The following tables show the expected performance at each current based on various manufacturer's specifications. In each cell are two numbers. The lower number is from the best known specification. The higher number reflects the worst specification found. Many manufacturers specifications were not very complete. In this case, numbers were used which were considered optimistic for the lower number and pessimistic for the higher. Table 4 in Appendix B shows the specifications used. Variation in output signal level was compensated for by the addition of hypothetical amplifiers to the low output voltage devices. These amplifiers are considered to have optimistic performance with cost limited to about a thousand dollars. Specifications for these amplifiers are shown in Table 5, Appendix B.

Conclusion

Precision current regulation requires precise current measurement. The degree of precision required, the maximum current and the operating environment govern the choice of a current measurement device.

The examined devices exhibit a variety of characteristics. The dominant factor affecting the precision of a current measurement device is it's sensitivity to temperature. Long term stability, repeatability, linearity, line voltage sensitivity, leakage flux sensitivity and signal level will also affect the choice of a current measurement device.

Acknowledgments

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Current Measurement Device	Long Term Stability (mA/month)		Variation Over Temp and Line Voltage Range (mA)		Repeatablity Over Power Cycles (mA)		Sum of Low Numbers "Precision" (mA)
	Low	High	Low	High	Low	High	Total Low
50mV Manganin shunt with ideal amp	0.3	0.6	113	150	0.3	0.6	113
50mV Manganin shunt with preamp	12	31	270	615	30	61	313
50mV shunt with preamped iso amp	12	31	349	810	30	61	392
50mV Manganin shunt with iso amp	120	301	7688	15338.	120	601	7928
200mV Hall current sensor with preamp	303	151	1170	37763	1500	6000	2973
Hall effect DCCT with ideal shunt	300	150	788	3938	600	3000	1688
Transductor	30	300	188	5250	30	1500	248
2nd Harmonic DCCT	3.3	10	45	165	3.0	15	51
2nd H. DCCT with temp control	1.7	6.6	4.5	41	3.0	15	9

Table 2: Performance of 3,000 amp current measurement devices at 300 amps as described above.

Current Measurement Device	Long Term Stability (mA/month)		Variation Over Temp and Line Voltage Range (mA)		Repeatablity Over Power Cycles (mA)		Sum of Low Numbers "Precision" (mA)
	Low	High	Low	High	Low	High	Total Low
50mV Manganin shunt with ideal amp	3.0	6.0	1125	1500	3.0	6.0	1131
50mV Manganin shunt with preamp	15	36	1350	2100	33	66	1398
50mV shunt with preamped iso amp	15	36	2138	4050	33	67	2186
50mV Manganin shunt with iso amp	123	306	9375	18375	123	606	9621
200mV Hall current sensor with preamp	303	151	4613	20651	1500	6000	6416
Hall effect DCCT with ideal shunt	300	150	1125	6525	600	3000	2025
Transductor	30	300	525	18750	30	1500	585
2nd Harmonic DCCT	6.0	18	113	300	3.0	15	122
2nd H. DCCT with temp control	3.0	12.0	11.3	75	3.0	15	17

Table 3: Performance of 3,000 amp current measurement devices at 3,000 amps as described above.

Appendix A
Definitions

Specifications for current measurement devices often include confusing and nonstandard terms. The following definitions have been written for purposes of this paper's clarity, and should work well in many power supply applications. These definitions are consistent with the IEEE Standard Dictionary of Electrical and Electronic Terms [1]. For simplicity, only DC characteristics will be discussed, unless otherwise noted.

Accuracy is the quality of freedom from error. Accuracy is expressed by the ratio of the indicated value to the true value. It is usually expressed in parts per million (ppm) or as a percent (1 ppm = 0.0001%). Since the true value cannot be determined exactly, the measured value or calculated value of highest available accuracy is taken to be the true value. Hence, when a device is calibrated in a given echelon, a comparison to a standard calibrated at a higher echelon is used to determine the true value.

Resolution is the smallest change in a signal which can unambiguously be discerned. Resolution of an analog signal is limited by noise, either internal or external. Resolution of a DC analog signal depends on the bandwidth of the signal. The signal can always be filtered more to produce higher resolution. Resolution does not imply accuracy.

Repeatability is the closeness of agreement of measurements of a device under exactly the same conditions after a period in time in which one or more of the conditions have been changed. A specification for repeatability should include measurement conditions and specific changes for the conditions.

Precision is the sum of repeatability and resolution. Precision does not imply accuracy. It does, however, imply that a device is capable to being calibrated of an accuracy class consistent with its precision. It is always the case that a device is more precise than it is accurate.

Stability has to do with the ability of a system to reach a steady state after a perturbation. It only affects precision in determining the time needed to obtain a good measurement after a perturbation.

Long Term Stability is the ability of a device to hold its true value over a long period of time under constant environmental conditions. For a current measurement device this means the current, temperature and power, etc., remain constant. The period of time depends on the application, it can be from hours to years. Note that long term stability does not include repeatability.

Calibration Interval is the time between successive calibrations. Calibration interval depends on the application and the needed accuracy. It is determined from the long term stability and repeatability of a device.

Transfer Impedance (Zt) can be defined for a two port device as the voltage at port two, divided by the current at port one, while the current at port two is zero. Zt is usually expressed in ohms. Transfer impedance for a current measurement device is sometimes referred to as gain.

Transfer Impedance Error or gain error is the difference between the actual Zt and the nominal Zt divided by the nominal Zt. Zt error is usually expressed in ppm or percent.

Relative Offset Error (offset error) can be the output voltage while the current in both ports is zero, divided by the full scale (F.S.) output voltage. Assuming a linear transfer impedance, the offset error can also be thought of as the current input needed to bring the output to zero, divided by the full scale current. Relative Offset Error is expressed in ppm of F.S. or percent of F.S..

Temperature Coefficient of Transfer Impedance. Temperature coefficient (T.C.) of transfer impedance is defined as the change in Zt due to a change in temperature, divided by that change in temperature. T. C. of Zt is usually specified in ppm/°C over some temperature range and is usually calculated by making measurements only at the end points. In general, a plot of the change in transfer impedance of a current measurement device versus temperature is not a straight line. Thus, any temperature coefficient specification should also include a temperature range or a specific temperature of interest.

Temperature Coefficient of Offset Error. T. C. of offset error is defined as the change in offset error due to a change in temperature, divided by the change in temperature. T.C. of offset is not the same at all temperatures. A complete specification should include the temperature of concern.

Linearity Error is the maximum deviation from a best fit straight line, which does not necessarily go through zero. Be careful if linearity is important! It is expressed in many different ways and is not comparable from device to device. Specifications for linearity give numbers as peak to peak, peak, or RMS differences. These numbers are sometimes specified as ppm of full scale or ppm of reading. The specifications many times do not explain which is being used. The relationship between each method depends on the shape of the nonlinearity. In addition, the nonlinearity can be time dependent. For example, a shunt is perfectly linear if the current is in it for a short time. Only when a shunt is allowed to warm up is it nonlinear.

Leakage Flux is caused by nonuniform magnetic fields in a measurement core [4]. Consider the magnetic field in a toroidal current transformer with a magnetic core of finite permeability. At exact ampere turn balance the flux in the core will be zero only if the windings are exactly symmetrical about the core and the core is perfectly round. If the current transformer is in some way asymmetrical, the flux would not be zero everywhere in the core. This flux is leakage flux. Leakage flux can be caused by any external distortion of the field such as that created by neighboring currents or magnetic materials.

Appendix B
Data Used for Comparison

Device	Long Term Stability of Offset (ppm of FS/month)		Long Term Stability of Gain (ppm/month)		Temperture Coefficient of Offset (ppm of FS/°C)		Temperture Coefficient of Gain (ppm/°C)		Repeatablity of Offset (ppm)	
	High	Low	High	Low	High	Low	High	Low	High	Low
Manganin shunt	0	0	1	2	0	0	15	20	1	2
200mV Hall current sensor	100	500	0	0	10	250	50	250	500	2000
Hall effect DCCT	100	500	0	0	10	50	5	25	200	1000
Transductor	10	100	0	0	2	50	5	200	10	500
2nd Harmonic DCCT	1	3	1	3	0.5	2	1	2	1	5
2nd H. DCCT with temp control	0.5	2	0.5	2	0.05	0.5	0.1	0.5	1	5

Table 4: Specifications used for the comparisons presented in Tables 2 and 3.

The amplifiers in Table 5 were used in the comparison section to help illustrate the problems with low output voltage devices.

Specification	Pre amplifier	Isolation amplifier	Units
Offset T.C.	0.1-0.3	5-10	uV/ C
Gain T.C.	1-2	10-25	ppm/ C
L. T. Stability	0.2-0.5	2-5	uV/month
Repeatability	0.5-1	2-10	uV

Table 5: Hypothetical amplifiers used for comparison purposes

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