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Improved Design Liquid Helium Temperature
Regulator Using Operational Amplifier Circuits

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ABSTRACT

We describe a cryogenic temperature regulator of unusually simple design, operation, and construction utilizing miniature solid-state operational amplifiers. The combination of a unique saturating amplifier and an analog multiplier type of cross-correlation phase detection circuit eliminates fast-risetime signals and switching transients, enabling the regulator to be used in a low-noise shielded environment. The circuit is of the conventional ac Wheatstone bridge type using a carbon resistance thermometer for the sensing element and providing a proportional heater feedback of > 1 W to the bath. The balanced input circuitry has been optimized for thermometer resistances between $500\ \Omega$ and $5\ \text{k}\Omega$, and the regulator is capable

of detecting a temperature change of 30 μ K at a thermometer power dissipation of $< 3 \times 10^{-11}$ W. Regulation of a superfluid 4 He bath to better than $\pm 4 \mu$ K for periods greater than 30 minutes has been achieved at a thermometer power of $< 5 \times 10^{-9}$ W with a response time of 0.1 sec. Owing to the high gain and broad bandwidth of this regulator, it is adaptable to a wide variety of other situations, such as a sample mounted on a copper block outside the bath, without encountering the usual problems of excessive power dissipation, instability, and thermal oscillations.

INTRODUCTION

Wheatstone bridge circuits for the measurement and control of temperature in the liquid ⁴He region are currently in wide use.¹⁻⁶ Such units utilize the large temperature coefficient of resistance of a small carbon resistor⁷ used as one arm of the bridge by amplifying the off-null signal and using it to control the power delivered to a heater placed in good thermal contact with the system to be regulated. The necessity for low power dissipation^{8,9} in the thermometer (requiring that the associated circuitry have very high gain), together with the desirability of a fast response to temperature fluctuations, indicates the need for a fairly sophisticated low-noise circuit. Although commercially available instruments can readily be adapted to this usage their cost tends to be prohibitively high for most users, particularly if it is desirable to have several instruments in use simultaneously. Accumulated experience with a number of different home-made bridges constructed of discrete components indicates that they tend to be complex, expensive, and time-consuming to construct, tricky to adjust, and difficult to repair quickly when they malfunction. To avoid these problems, we have designed an extremely simple, easy to construct and maintain, and relatively inexpensive regulator based primarily on commercially available solid-state operational amplifiers and having a minimum of discrete parts. To facilitate construction, isolation, and repair and to allow quick interchangeability in critical situations, the unit was constructed in the form of four plug-in modules in a miniature NIM type bin.

The first two sections of the paper discuss the design and construction of the regulator circuitry. The third section deals with the important stability criteria, while the following two sections cover the operation and performance of the completed regulator. As a result of the rapid development, steadily increasing performance, and rapidly diminishing prices of miniature solid-state devices, we recommend that current price lists and catalogues be consulted before embarking on construction; the units we chose will most probably be obsolete by the time this appears in print. Owing to the operational amplifier circuit design used, newer or more efficient devices may be substituted freely without affecting circuit performance. The overall success of this design is illustrated by the large number of units presently being constructed in our department.

DESIGN CONSIDERATIONS

The usual considerations in the design of a regulator of this type are: a) the necessity for low power dissipation in the thermometer; b) the short-term deviation and long-term drift of the temperature and; c) the overall stability of the system, i.e., freedom from regenerative feedback or "oscillation". The latter is a severe problem when attempting to regulate ⁴He bath temperatures between 4.2°K and the λ -point in many of the available regulator designs. In addition, two rather more stringent conditions were added to our considerations. First, it was desirable to keep the thermometer power dissipation even lower than usual so that we could measure temperatures in the liquid ³He range, or in experiments where the sample was only weakly connected to the bath, without undue

error¹⁰ or excessive heat input. Secondly, it was desired to operate one or more of these units in a screened or shielded environment, which imposed the condition that there be no transients or fast-switching circuits under any operating conditions. An effort was also made to keep the regulator as simple and straightforward to operate as possible, with a minimum number of controls to adjust.

CIRCUIT DESIGN AND CONSTRUCTION

General

A block diagram of the unit is shown in Fig. 1. The broken lines indicate the grouping of the circuitry into three distinct modules for purposes of isolation. The entire unit is powered by a single, tracking, ± 15 Vdc power supply built into a fourth module. The entire four module assembly occupies only 17.8 cm of panel space in a conventional 48.3 cm relay rack and could easily be made more compact.

The oscillator frequency was chosen to be 33 Hz to keep the fundamental and its harmonics well below frequencies used for experimental purposes, far from 60 Hz, and still high enough to avoid f^{-1} noise in the amplifiers. The oscillator is used both to drive the bridge and, via a phase shifter, to supply a reference signal to the detector.

The error signal from the ac Wheatstone bridge is amplified by a low-frequency transformer, a low-noise preamplifier which includes a 60 Hz notch filter, a tuned amplifier, and a saturating amplifier of unique design; at this point the small signal voltage gain is over 140 db. The error signal is then multiplied with a properly phased reference signal and the second harmonic component of the resultant filtered out by a 66 Hz twin-tee. The resulting signal is time averaged by a low pass

filter and used to drive the heater via a high current amplifier. At optimal thermometer values and heater settings an overall power gain of over 150 db is achieved, giving closed loop gains as high as 10^4 W/K.³

Bridge and Preamplifier

The ac Wheatstone bridge and the preamplifier are constructed in a single module as shown in Fig. 2. The value of the fixed resistors was chosen to keep the power dissipated in the thermometer roughly constant for resistance values between 1 k Ω and 100 k Ω when the bridge is driven by a low impedance oscillator. Figure 3 displays a set of power dissipation vs. resistance curves for the thermometer resistor as a function of the setting of the oscillator level switch. A three-wire configuration was chosen for the external connection to the thermometer to minimize ac pickup effects and cancel out the resistance of the leads; the input common lead and the variable bridge arm lead should be connected together at the thermometer. Owing to the low input impedance of the bridge and the balanced cable configuration any reasonable length of input cable may be used without degrading the regulator performance. The bridge is balanced with a six-dial miniature precision decade resistor.¹¹ Lead reactances are compensated by means of the adjustable capacitors shunting the two variable arms of the bridge.

The error signal is first amplified by a very well shielded low-frequency input transformer having a gain of 37.7:1.¹² The physical placement of this transformer is critical due to the low signal levels involved and should be adjusted experimentally to a position which minimizes pickup from adjacent circuitry. Three benefits are conferred upon us by the use of a transformer input: the input impedance is lowered,

enabling long cable runs; the signal-to-noise ratio is enhanced by the transformer gain, and; the noise-limited resolution of the bridge is optimized in the region of 500 Ω to 5 k Ω thermometer resistances where we usually operate. The first stage X50 amplifier has an FET input to raise the input impedance of the transformer as high as possible. This operational amplifier must, in addition, have low input noise, high CMRR, and an extremely high tolerance to power supply fluctuations to prevent any 33 Hz present in the power supply leads from appearing at the input. The unit chosen¹³ has proved so effective that, providing the power supply leads are properly routed and the common grounded only to the center of the bridge as shown on the schematic, the entire unit (including the oscillator) may be driven by a common power supply without difficulty. Any line-frequency pickup present is then removed by the 60 Hz twin-tee, and the signal is further amplified by a buffer amplifier¹⁴ used to provide a low output impedance without loading the twin-tee network. For convenience of operation we chose the buffer gain to be X4, which limits the meter jitter due to inherent noise in a 10 Hz bandwidth to about 5% of full scale for thermometer resistances below 10 k Ω . The gain of the buffer amplifier may be increased simply by altering the ratio of the feedback resistors if additional overall gain is required.

Correlator and Heater Driver

The second module contains the balance of the regulation circuitry as shown in Fig. 4. Because of the excellent overload performance designed into the circuitry, the 5 k Ω semilogarithmic potentiometer is the only gain control necessary; it is used primarily to facilitate coarse balancing of the bridge. The signal is first amplified by a

double-tuned, twin-tee, transistorized active filter¹⁵ having a center gain of X100 and a flat-topped bandpass characteristic with a 23% bandwidth. This is the only tuned amplifier in the circuit, and its passband was deliberately chosen to be quite wide to insure fast response with minimum phase shift near the center of the band. The 0.47 μ f capacitor was inserted to block the small dc voltage present at the output of the tuned amplifier and is not otherwise necessary or desirable.

Following the tuned amplifier is a saturating amplifier of unique design. Basically, it is a bipolar logarithmic amplifier with the feedback diodes shunted to limit the small signal gain. The feedback signal is taken from a voltage divider at the output to offset the logarithmic portion of the characteristic. Figure 5 illustrates the resulting transfer characteristic of the circuit ("input voltage" refers to the signal at the output terminal of the tuned amplifier). The gain remains constant at X40 over the center $\pm 25\%$ of the output range and then falls off rapidly, the output voltage essentially saturating at 6.5 V for the 8V to 10V maximum input signal available. It may also be seen from Fig. 5. that the output waveform remains "soft", allowing the amplifier to clip the input waveform without generating significant power at high frequencies or inducing fast rise transients. The resulting gain compression of about 70:1 at 33 Hz enables the bridge to be balanced quickly and easily without excessive manipulation of the gain control, while maintaining a constant maximum sensitivity near null. Although the feedback equation for this circuit is not solvable in closed form owing to the non-linear response of the diodes, it may be solved graphically using the equivalent circuit shown in Fig. 6. As

the inverting input terminal is a virtual ground which draws no current because of its high input impedance, the input voltage and source impedance may be replaced with an ideal current generator whose current is given by $i_g = e_{in}/|Z_{in}|$, or

$$i_g = e_{in}/13 \text{ k}\Omega. \quad (1)$$

The important parameter in the equivalent circuit is e_x , since the output circuitry must drive just enough current through the $100 \text{ }\Omega$ resistor to maintain e_x at a potential equal to the voltage drop induced by i_g across the $50 \text{ k}\Omega$ resistor and its diode shunt. Assuming the resistance values shown and an ideal exponential characteristic for the 1N483B diode¹⁶, the resulting pair of simultaneous equations are,

$$e_{out} = 10.35 e_x + 0.072 e_{in}, \quad (2)$$

and

$$e_{in} = 0.26 e_x + 1.3 \exp(23 e_x - 11.5). \quad (3)$$

The solution to this pair of equations is shown graphically in Fig. 5. The overall small signal gain of the regulator measured at the output of the saturating amplifier is greater than 140 db.

Having thus sufficiently amplified the signal the next step is to demodulate it, and some form of phase sensitive detection is the most desirable way of doing this. Conventional phase detection circuits using half-wave or full-wave switching, whether mechanical or electronic, were felt to be undesirable for two reasons. As previously mentioned, every effort was being made to avoid any switching or fast rise circuitry

at all in order to eliminate noise-producing transients. Also, most phase-sensitive detector circuits have, superimposed on the desired dc output signal, a signal of comparable amplitude at the fundamental frequency. This requires the use of a low-pass output filter which either has a cutoff point well below the fundamental frequency, or a very steep attenuation slope above cutoff: The first of these would limit the frequency response too severely (since the bridge operates at 33 Hz) and the second would yield a very undesirable equivalent gain-frequency plot for the regulator, leading to possible unstable operation under some conditions. For these reasons, we chose to demodulate the signal by cross-correlating¹⁷ the amplified error signal with a properly phased reference signal derived from the bridge oscillator. A small analog multiplier¹⁸ is used to perform this multiplication.

The cross-correlation method of demodulation is based on the fact that the time average of the product of the error and reference signals as a function of their relative phase is equivalent to the usual cross-correlation between two sinusoidal signals,

$$R_{12}(\tau) \equiv \lim_{T \rightarrow \infty} \frac{1}{2T} \int_{-T}^T [A_1 \sin \omega_1 t] [A_2 \sin \omega_2 (t + \tau)] dt. \quad (4)$$

If we set $\omega_1 = \omega_2 \equiv \omega$ and replace $\omega\tau$ with the phase difference, ϕ , this reduces to

$$R_{12}(\phi) \equiv \frac{A_1 A_2}{2} \left\{ \lim_{T \rightarrow \infty} \frac{1}{2T} \int_{-T}^T [(1 - \cos 2\omega t) \cos \phi - \sin 2\omega t \sin \phi] dt \right\}. \quad (5)$$

For times $T \gg \omega^{-1}$ this yields

$$\langle R_{12}(\phi) \rangle_{av} = \frac{1}{2} A_1 A_2 \cos \phi \quad (6)$$

which proves that the cross-correlation method is equivalent to other methods of phase sensitive detection. Looking back to Eq. (5), however, we note that the effect of inserting a 66 Hz twin-tee at the multiplier output is to remove the terms in $2\omega t$ from the integrand. In this case, the unaveraged correlation function is given by

$$R_{12}(\phi) = \frac{1}{2} A_1 A_2 \cos \phi \left\{ \lim_{T \rightarrow \infty} \frac{1}{2T} \int_{-T}^T dt \right\} \quad (7)$$

which reduces to Eq. (6) for any finite time constant. In principle no averaging is necessary, and the response time of the regulator should be limited only by the characteristics of the twin-tee notch filter. As this filter has a 66 Hz null with a very sharp notch, response times of the order of the reciprocal of the fundamental frequency are attainable, and by operating at a higher frequency than the 33 Hz used in the present design extremely fast response could be achieved. In practice, however, the error signal also has a large noise power spectrum, which results in a finite ac output signal from the multiplier over a broad frequency range. The integrator following the 66 Hz twin-tee is used as a low-pass filter to attenuate this noise. (Any harmonics of the 33 Hz drive frequency present on the error signal will be multiplied up to higher frequencies and the filter will also remove these.) Since there are no large amplitude coherent signals remaining below 99 Hz,¹⁹ a relatively short time constant of 0.1 second is used, with a simple RC rolloff of 6 db/octave. The resulting dc output is read by a small zero-center panel meter.

The heater drive circuit consists of another operational amplifier which has a current booster to drive the heater from the + 15 volt

supply. As the heater current must be controlled about some finite mean value, the inverting input terminal is used as a summing point for the dc error signal and a variable offset voltage derived from the -15 volt supply line. With the mode switch in manual, the 10 turn potentiometer is adjusted to provide a finite heater current. When the switch is placed in automatic the error signal is summed with the bias signal, and the output current of the amplifier will vary about the manual value accordingly. The inverting input being a virtual ground, there is no interaction between the error and bias circuits, nor are there any transient effects when the manual-auto switch is actuated in either direction. An npn Darlington pair is used as a booster to increase the output current; the high input impedance of such a pair avoids loading the operational amplifier, while the very low output impedance facilitates driving the heater. Since the voltage gain of the pair is close to unity the closed loop gain is not affected. A 7V Zener diode is inserted in the output loop to insure that no value of the input signal will drive the output current through zero into the reverse direction (which would obviously lead to positive heater-thermometer feedback and wild temperature oscillations). An analysis of the heater circuit shows that there are actually two feedback loops around the amplifier. The error signal loop consists of a 6.8 k Ω input resistor and a feedback resistor which varies from 22 k Ω to 24 k Ω , depending on the manual heater setting; the error signal loop gain is thus about 3:1. In the manual heater setting loop, the 22 k Ω resistor draws no current and becomes part of the input circuit; the "gain" of the amplifier (using the -15 volt supply as a signal) is controlled by the manual heater

adjust potentiometer. This "gain" may be varied from zero to 0.883:1 giving a maximum of 12.5 volts across the heater at currents up to about 120 mA. A heater winding of 120 Ω or greater is recommended to avoid overloading either the + 15 volt supply or the drive transistor. Although the resulting maximum heater power of > 1 watt should be adequate for most applications, the maximum heater current may be increased very easily by providing a more robust power supply and changing the output transistor accordingly.

Oscillator and Phase Shifter

The third module, containing the oscillator and phase shifter circuitry, is diagrammed in Fig. 7. The oscillator itself is a conventional twin-tee feedback design with amplitude control provided by a diode limiter circuit. A 5 k Ω adjustable resistor is used to complete the tee, and should be set to a value which just sustains stable oscillation. A 10 k Ω potentiometer is included in the amplitude limiting loop to enable the output amplitude to be adjusted to the specified value. The amplifier immediately following the oscillator drives both the bridge and one side of the adjustable phase shift network. It is followed by a similar stage whose output is 180° out of phase with the first to drive the other side of the adjustable phase shifter. The constant amplitude phase shifting network shown is continuously adjustable from 0° to 180° by adjustment of a single 5 M Ω potentiometer.²⁰ A true logarithmic taper is used to give a more linear relationship between the control setting and the phase; for true linearity a custom-made potentiometer would be necessary. This network, used in conjunction with the phase switch at the bridge drive output, provides a full 360° phase range. The buffer amplifier used to

isolate the reference drive output has an FET input stage to avoid loading the adjustable phase circuit. It is absolutely essential to the proper operation of the correlator that no dc output be present on the reference signal;¹⁹ if the amplifier used has no dc zero adjustment, a coupling capacitor may be required.

This circuit has been grossly overdesigned so as to make it easily adaptable to a variety of other uses in our laboratory. It can readily be simplified by using a conventional transistor phase splitter to drive the phase shift network and an ordinary FET to drive the reference output. This will reduce the cost considerably.

Power Supply

As previously mentioned, all the dc power needed to supply the instrument is obtained from a dual ± 15 Vdc @ 300 mA power supply contained in a fourth plug-in module.²¹ The requirements for good regulation, very low output impedance and good tracking between the +15 V and -15 V lines can also be met by a large number of commercially available supplies. The following precautions in connecting the supply are necessary for proper operation. The common line from the supply must be grounded only at the center of the ac Wheatstone bridge, via a lug on the air variable capacitor. From this point, the common is run through the preamp following the path of the error signal. The other modules also have internal common lines which connect to the power supply common at only one point. The ± 15 V lines should connect first to the input amplifier and then follow the signal path through the unit, connecting finally to the oscillator stages. In addition, the extremely high sensitivity of the input stages requires that the power supply be located

physically as far away from the preamplifier-bridge module as possible. The above precautions, together with the very high (over 100 db) power supply offset rejection of the input operational amplifier, allow the unit to operate easily off a single supply. As the total current drain of the preamplifier module is small (typically it will be less than 25 mA) battery operation is possible; this would also allow for remote operation of the bridge module if desired.

STABILITY

A major problem in any servo-loop regulator is stabilization of the system against oscillation over a wide variety of operating conditions. The simplest method for analyzing such systems is to treat the electrical part of the servo control as a single operational amplifier, using the external parameters to determine the feedback conditions. In this approximation we replace all of the ac electronics of the regulator by a single equivalent chopper-stabilized operational amplifier having a dc open loop gain of 150 db. The Bode (gain vs. frequency) plot of the equivalent amplifier depends on the time constant chosen, and is further complicated by the presence of the two twin-tee notch filters. Ignoring the effect of these tees, we can construct a simplified Bode plot for the several time constant settings. For an 0.1 sec time constant, the gain is flat at 150 db out to 3 Hz, 12 db down at 10 Hz, and rolls off roughly 12 db/octave (40 db/decade) at higher frequencies, ultimately reaching unity gain at about 15 kHz. The stability of the regulator in any closed loop configuration is governed by the difference in slope between the open loop and closed loop gain-frequency characteristics at the point where they intersect. If this intersection occurs at a gain greater

than unity, the regulator will be conditionally stable for a difference in slope of < 40 db/decade and absolutely stable for a difference of < 20 db/decade.

Assuming a very tight heater-thermometer coupling, such as that achieved by immersion in a superfluid helium bath, the closed loop servo response is closely approximated by a voltage follower circuit, whose gain is roughly unity, and the regulator will be stable at virtually any control setting. In any more general experimental set-up (such as immersion in liquid ^4He above the λ -point) the weaker coupling between the heater and the thermometer will allow the equivalent closed loop gain to be greater than unity. The stability of the system will then depend on whether or not the closed loop and open loop gain vs. frequency characteristics intersect at greater than unit gain. If we also assume the presence of an appreciable thermal mass in the system, this mass acts as an integrating capacitor in parallel with the direct feedback, providing a high frequency rolloff for the closed loop response. By keeping the open loop gain and bandwidth as large as practicable we have attempted to insure that, under the widest possible operating conditions, either the closed and open loop characteristics will intersect at less than unit gain or the closed loop response will be falling off rapidly enough at the intersection to give a rate of closure less than 20 db/octave. If an instability should occur, the simplest cure to try would be to replace the double-tuned 23% bandwidth amplifier stage with a similar unit having either a greater bandwidth or a lower Q response. This will degrade the regulation less than the alternative choice of increasing the time constant of the output filter.

OPERATION

Initial alignment of the regulator consists of adjusting the oscillator for best operation and correct amplitude, and dc zeroing the operational amplifiers. Once set, these controls should require readjustment infrequently, if at all. Most analog multipliers have a recommended external null suppression or offset trimmer resistor which should be adjusted to minimize the amount of reference signal fed through to the output with the error signal input grounded.

With the thermometer connected and the bridge well off null, the phase of the oscillator should be adjusted to close the Lissajous figure displayed on any convenient oscilloscope via the X and Y output connectors. (In lieu of an oscilloscope this can be done using the panel meter alone by adjusting for maximum deflection.) If the resistance of the thermometer changes by an order of magnitude or more during a run, it will be necessary to readjust the phase setting. The 180° phase reversal switch position should be checked to make sure that the heater current is varying in the correct direction when the thermometer resistance changes!

In most applications the oscillator level is preset at the highest value allowed by maximum thermometer power dissipation considerations. This is greatly facilitated by the flat power-resistance curve, as shown in Fig. 3. The heater is first adjusted (in manual) to an empirically determined current setting which provides adequate heat input for effective control of the bath. A coarse null can readily be achieved at minimum gain and steadily perfected as the gain is increased. When a null is achieved at full gain, the phase shifts due to reactances

in the decade and thermometer resistors should be compensated by adjusting the variable capacitors so as to close the Lissajous figure into a straight horizontal line. This can also be done with the panel meter alone by checking that the null is not affected by rocking the phase control (the addition of a 90° phase shift switch to the oscillator would simplify this panel meter procedure enormously). After the bridge is completely nulled the heater switch is placed in automatic to commence regulation. If the temperature starts to drift in either direction, the null balance heater current should be adjusted with the manual adjust control.²³ Once locked on, the regulator is completely stable and may be left unattended for long periods of time.

Owing to its excellent overall stability, the regulator may also be used to control the temperature over a limited range merely by altering the decade resistor setting. The temperature will adjust to the new value and settle down almost immediately, allowing one to preset a desired temperature with high accuracy. Other features of our circuit design include quick, stable recovery from transient heat pulses and freedom from long-term drift.

PERFORMANCE

We believe that any general statement of regulator performance is misleading, as the regulation depends critically on such parameters as the bath temperature and volume, the thermometer power, and the null balance heater current, all of which must be adjusted to balance maximum possible sensitivity against experimental limitations. A much more useful method of predicting regulating ability is the use of Fig. 3 as a regulator nomogram. In Fig. 3 we plot, in addition to

thermometer power dissipation curves, experimentally determined curves of constant resolution, $\Delta R/R = 10^{-3}$ and $\Delta R/R = 10^{-4}$, taken at a signal/noise ratio of unity and a time constant of 0.1 seconds.²⁴ Assuming that sufficient heater power is available and allowable, and knowing the temperature coefficient and actual value of the thermometer at the desired temperature, the maximum temperature resolution is easily determined.

Carbon resistors ordinarily used for cryogenic thermometers have a temperature sensitivity, $d(\ln R)/dT$ on the order of 2 K^{-1} to 6 K^{-1} at their operating temperature if their value is carefully chosen. Translating the nomogram in these terms, we observe that the predicted maximum regulating ability using such a thermometer, at a temperature corresponding to a resistance value of between 500Ω and $4 \text{ k}\Omega$, will be $\approx 30 \mu\text{K}$ with a power dissipation of $\leq 5 \times 10^{-11} \text{ W}$. This can be improved to $\approx 3 \mu\text{K}$ by increasing the thermometer power 20 db to $\leq 5 \times 10^{-9} \text{ W}$. In a specific application in our laboratory, using a 27Ω nominal $1/8 \text{ W}$ Allen-Bradley resistor, a heater consisting of $\approx 200 \text{ cm}$ of bifilar wound resistance wire distributed evenly over a 20 cm long copper can 5 cm in diameter which enclosed the entire experiment, and an 8.5 liter glass liquid helium dewar, we were able to maintain the sample temperature to within $\pm 40 \mu\text{K}$ for several hours without adjustment. The set temperature was 1.270 K (corresponding to a thermometer resistance of $8.8 \text{ k}\Omega$), the thermometer dissipation was $9 \times 10^{-11} \text{ W}$, and it was necessary to supply 20 mW of average heater power to maintain control. By raising the thermometer power to $9 \times 10^{-9} \text{ W}$ the regulation was improved to $\pm 5 \mu\text{K}$, but it was necessary to readjust the manual heater control every 30 minutes or so to compensate for drift. The above performance compares

very well with the nomogram predictions for a 10 Hz bandwidth. The quoted performance could be improved by increasing the buffer amplifier gain and using a longer time constant; the latter, however, is undesirable in our experimental setup because of the low heat capacity of our apparatus at these temperatures. The observed drift was primarily a result of the use of an unregulated pumping system; it could be further reduced by increasing the heater power dissipation, at the expense of higher ⁴He consumption, or regulating the bath pressure.

If increased temperature resolution or lower heater power is necessary, the buffer amplifier gain should be increased by a factor of three or more and the time constant set at 1 sec. This increases the resolution at constant power by a factor of three, and lowers the power at constant resolution by a factor of ten. The nomogram can still be used by either shifting the constant resolution profiles -10db or relabelling them. Under these conditions a resolution of $< 1 \mu K$ can be realized at a dissipation of $< 3 \times 10^{-9} W$.

FURTHER REMARKS

The regulators currently in use in our laboratory are constructed of the discrete circuit type of packaged miniature operational amplifiers. This is convenient for semi-experimental circuit work but somewhat expensive. The use of integrated circuit operational amplifiers would bring the cost down appreciably and lower the quiescent power consumption. Great care should be exercised in choosing the input amplifier, which should have low noise and a very high CMRR as well as a high degree of immunity to power supply fluctuations. The overall size of the complete unit can be reduced to 25 x 15 x 15 cm, but any further reduction is

limited by the necessity of allowing enough finger room to manipulate the controls and would also introduce serious shielding problems. The overall parts cost was about \$800 per unit: This can be decreased about \$200 by using integrated circuit operational amplifiers and simplifying the oscillator module; any further reduction in cost would have to be at the expense of replacing high quality components such as the Dekastat with less expensive equivalents.

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FIGURE CAPTIONS

- Fig. 1. Block diagram of the temperature regulator. The broken lines indicate the grouping of circuit elements into three plug-in modules. The power supply module is not shown.
- Fig. 2. Circuit diagram of the Bridge and Preamplifier module.
- Fig. 3. Thermometer power dissipation vs. resistance nomogram. The solid lines refer to the six settings of the oscillator level switch. The broken lines indicate experimentally determined constant resolution profiles which are compared to the theoretical noise-limited performance (straight horizontal lines). For further details see Footnote 24.
- Fig. 4. Circuit diagram of the Correlator and Heater Driver module.
- Fig. 5. Transfer characteristic of the saturating amplifier. Input voltage is measured at the output terminal of the preceding tuned amplifier. The inset shows the saturation waveform at the maximum input amplitude of 10 V peak.
- Fig. 6. Saturating amplifier current equivalent circuit. The input terminal is at virtual ground as indicated by the broken line.
- Fig. 7. Circuit diagram of the Oscillator and Phase Shifter module.

FOOTNOTES

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10. Using the formula $dW/dT = 4 \times 10^{-5} T^{1.6}$ (W/K) of Ref. 9, the maximum allowable thermometer power for 100 μ K accuracy at 0.3 K is 6×10^{-10} W. Using a Speer type 1002 470 Ω nominal resistor our unit can resolve 100 μ K at 0.3K with a dissipation of only 3×10^{-10} W.
11. Electro-Scientific Instruments Model DS625A Dekastat.
12. Triad G-10 Geoformer. A custom-made transformer having higher input and output impedances would improve the regulator performance.
13. A Burr-Brown Research Corp. model 3062/15 was used.
14. Except for the input amplifier, all FET operational amplifiers used were Burr-Brown model 3116/12C; transistor input types were Burr-Brown model 3118/12C. These may be replaced by similar units (see text).
15. White Instruments, Inc. type 273-23%; 33 Hz.
16. The characteristic of 1N483B diodes may be approximated by the formula $I = \exp(23V-20.7)$. Matched pairs were used to insure waveform symmetry.
17. The cross-correlation function, $R_{12}(\tau)$, is defined in most modern

books on stochastic processes. See, for example, A. Papoulis, Probability, Random Variables and Stochastic Processes (McGraw-Hill Book Co., New York, 1965), Chapter 9.

18. We have used several types manufactured by Analog Devices, Inc. and the Burr-Brown Research Corp. Several other companies manufacture similar units.
19. It is imperative that no dc voltage be present on either multiplier terminal, as this will produce a finite 33 Hz output signal.
20. Allen Bradley Type "J" composition potentiometer with a "DB" true logarithmic taper.
21. This power supply is one of a series designed and constructed by the electronics shop of the Dept. of Physics for powering modular instruments of this type.
22. Excellent descriptions of Bode plots and their uses may be found in the informative booklets published by such operational amplifier manufacturers as Philbrick-Nexus, Burr-Brown, and Analog Devices.
23. When working with ^4He above the λ -point, a mechanical pressure regulator of the diaphragm or manostat type will greatly minimize drift.
24. In Figure 3 the theoretical noise limit is given by the Johnson noise power, $4kT\Delta f$. For resistances $< 1 \text{ k}\Omega$ the resolution is limited by the amplifier voltage gain. The decrease in sensitivity at high resistance values is a consequence of the use of a low impedance detector transformer which loads the bridge. For thermometer resistance above $\approx 30 \text{ k}\Omega$ better performance is obtained by either replacing the transformer circuit with a balanced FET amplifier or using a higher input impedance transformer with less voltage gain.

Figure 1

XBL696-2899

Figure 2

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Figure 3

Figure 5

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Figure 6

Figure 7

XBL 696-2900

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