4. Conclusions
(i) A Langmuir–Blodgett trough with a diamond-shaped constant-perimeter barrier was reported. This had a number of advantages over other designs:
   (a) it was mechanically relatively simple and easy to construct,
   (b) the diamond was easy to remove and could be cleaned quickly,
   (c) there were no entrapment sites,
   (d) there were no surfaces in friction (rollers or sliding barriers) likely to generate particles within the active area of the trough,
   (e) the entire system was mounted in a laminar-flow enclosure (class 10), housed within a clean room (class 1000), to keep particulate contamination to a minimum.
   (ii) The trough produces excellent graphs of surface pressure against area per molecule; the latter is accurate to \(\pm 2\%\), the diamond area alone being accurate to better than \(\pm 1.5\%\).
   (iii) High-quality multilayer films may be deposited, free of gross ionic contamination and pinholes.
   (iv) The flow patterns during deposition are acceptable (figure 6) when compared with those reported for conventional trough designs (Daniel and Hart 1985, Daniel et al 1985).

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Figure 1. The bridge circuit.

The bridge circuit.

The circuit is centred around a Wheatstone bridge, one leg of which is the cold-wire probe. The upper two legs are 10 kΩ precision resistors (0.01%) and the last leg is a pair of wire-wound trim pots, 1 kΩ and 10 Ω, which serve as coarse and fine bridge balances. Current is fed to the bridge from a 6 V source regulated by a temperature-compensated Zener diode. The current is adjusted with a 100 kΩ variable resistor and an intermediate switch which provides 100 pA and 500 pA meter ranges. The load above the sensor is sufficiently large that fluctuations in the sensor resistance have a negligible effect on the current.

A meter reads either bridge current or output voltage. When it is switched out of the current line, an equal resistance is switched into that line in its place. A potentiometer sits in parallel with the meter to trim its scale.

The voltage difference across the bridge is delivered to the inputs of a commercial-grade Precision Monolithics OP-07 amplifier. This amplifier serves to buffer the output and provides the initial gain of the circuit. The amplifier is connected in a feedback configuration, summing at pin 1 rather than pin 2. This arrangement increases the input impedance at pin 2 so that the inputs are symmetrical with respect to the amplifier's internal circuitry. The common-mode rejection ratio is improved from 60 dB to more than 90 dB. Thus, the already low drift and noise levels of the amplifier are further enhanced. Such designs are discussed by Mortensen (1972) and Counts et al (1981).

The zero of the amplifier is adjusted by a variable resistor in conjunction with a reed relay zero check across the bridge. A pair of diodes across the inputs protect the amplifier. The output of this amplifier is sent to an additional pair of cascaded OP-07s for a final gain switchable between x1000 and x10000.

Temperature drift error in the bridge circuit is minimised by using metal film resistors and identical connecting wires wherever possible. The components of the bridge were selected to have nearly identical temperature coefficients of resistance, so that the resistors track together if the bridge temperature changes. Cermet trim pots, if used, should be used in conjunction with metal film resistors which provide the bulk of the load. Efforts should be made to use a single metal for wiring ahead of the amplifier inputs to minimise uncancelled thermoelectric potentials; such potentials may produce minor dc offsets. Good construction technique must be used throughout.

The bridge is powered with a remote supply to minimise heating effects and 60 cycle noise. The supply uses conventional full-wave rectification into three-terminal regulators at ±12 V. The supply signal must be free of 60 cycle noise. The bridge itself should be kept at a distance from unshielded sources of magnetic field to avoid induced 60 cycle noise.

3. Bridge response and performance

For fluctuations of sensor resistance small compared with the
Low-cost, high-performance DC cold-wire bridge

upper bridge resistor, the circuit's output is

\[ e = G I \Delta R, \]  

(1)

where \( R \) is the sensor resistance, \( \Delta R \) is the change from resistance at balance, \( G \) is the gain and \( I \) is the current through the sensor. The temperature dependence of sensor resistance is

\[ R = R_0 (1 + a_0 (\theta - \theta_0)) \]  

(2)

for moderate variation of temperature \( \theta \) (the subscript 0 denotes a reference state and \( a_0 \) is the temperature coefficient of resistivity). For such temperature changes, fluctuations in sensor resistance will be small enough that equation (1) holds and bridge output is related to the difference between sensor temperature and sensor temperature at balance by

\[ e = G I R_0 \Delta \theta. \]  

(3)

The temperature response will be limited by the frequency response and noise levels of the circuit. The frequency response of the circuit was determined by injecting a sine wave at the test input on the bridge and measuring the output response. The result, presented in figure 2, shows a -3 dB point of 8.2 kHz. In most cases, the frequency response of this circuit will exceed the frequency response of the sensor, the latter being less than 5 kHz for all but the smallest wires (see, for example, LaRue et al 1975). The output noise of the circuit was measured with a fixed load connected in place of the probe. The unfiltered noise is 13.6 mV RMS, and with low-pass filtering at 5 kHz the noise is 8.5 mV RMS.

The drift of the circuit was measured at a gain of 10 000 with metal-film resistor in place of the probe. The output was found to vary in a range of ±10 mV at a rate not exceeding 4 mV h⁻¹, and the drift was not correlated with temperature shifts over the range of temperature fluctuations encountered in our laboratory (±2 K). The bridge output saturates at +11.2 V and -10.5 V from which the operating range can be calculated using equation (3).

4. Spectral characteristics

To examine the spectral characteristics of the circuit, measurements were made in a turbulent grid flow with strong thermal stratification as illustrated in figure 3 (see also Lienhard 1988). The mean flow speed was 2.4 m s⁻¹ and the RMS turbulence intensity was 3% at the location of measurement. The cold-wire probe was a Dantec model 55P31 (1 μm diameter by 0.4 mm long platinum wire, -65 Ω) driven at 230 μA.

Typical temperature power spectra are shown in figure 4. The upper curve is the temperature spectrum measured with thermal stratification (RMS θ of 1.93 K) and has classical form. The next curve down is the spectrum of the same flow without heating. The RMS temperature measured in this case was 0.011 K. The isothermal curve is 55 dB below the heated curve at lower frequencies and shows line noise at 60 cycles and its odd harmonics.

The 'isothermal' spectrum possesses enough structure that it is worthwhile to examine the nature of the output signal for this case. The probe was replaced with a metal-film resistor and the bridge output measured; this voltage was converted to an equivalent temperature noise for the 55P31 sensor using

\[ \text{The equivalent temperature noise depends on the sensor current and resistance (diameter), with the lowest noise for higher currents and smaller sensor wire diameters. The objective of minimising noise must be balanced against the increased velocity sensitivity of the cold wire at higher currents and the greater fragility of smaller diameter sensors.} \]
equation (3). The associated temperature spectrum is shown in figure 4 \((\text{rms } \theta = 0.006 \text{ K})\), and coincides with the high-frequency portion of the isothermal spectrum. Thus, the high-frequency end of the isothermal spectrum represents the noise limit of bridge sensitivity. The Johnson noise of the sensor \((\text{rms } \theta = 7.5 \times 10^{-7} \text{ K})\) lies well below the observed noise limit, so the latter is purely electronic. The line noise spikes do not appear in the metal-film resistor spectrum and thus may be due to the electromagnetic pick-up of the cold wire itself.

To examine the low-frequency end of the isothermal signal, we may follow the general procedure of Wyngaard (1971) and compute the velocity-induced temperature signal. Wyngaard showed that

$$\theta - \theta^m = \frac{1}{\bar{H}} \frac{\partial \bar{H}}{\partial U} u \tag{4}$$

where \(\theta - \theta^m\) is the difference between actual and measured fluid temperature, \(\bar{H}\) is the time-averaged overall heat transfer coefficient for the wire, and \(u\) is the turbulent fluctuation in the mean velocity \(U\). To compute \(\bar{H}\) we use the equation of Nakai and Okazaki (1975):

$$\bar{H} = \frac{\pi l k}{0.8237 - \ln(Pe^{1/2})} \quad Pe = Ud/a < 0.2 \tag{5}$$

where \(l\) and \(d\) are the wire length and diameter, \(k\) and \(\alpha\) are the thermal conductivity and diffusivity of air, and \(Pe\) is the Péclet number based on diameter. Differentiating with respect to \(U\), we find

$$\theta - \theta^m = \frac{l^2 R}{2\pi l k U} u \tag{6}$$

where we have evaluated the coefficient for our flow.

Using equation (6) with the velocity spectrum measured in this flow, we obtain the velocity contamination spectrum shown in figure 4 \((\text{rms } \theta = 0.001 \text{ K})\). The contamination spectrum lies below both the isothermal and noise spectra. Apparently, the low-frequency structure of the isothermal temperature spectrum is a real temperature signal and cannot be attributed to velocity contamination.

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