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ARL-AERO-R-172

AR-004-509

DEPARTMENT OF DEFENCE

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AERONAUTICAL RESEARCH LABORATORIES
MELBOURNE, VICTORIA

Aerodynamics Report 172

A HYBRID HOT-WIRE DATA ACQUISITION SYSTEM (U)

by
J.H. Watmuff

Approved for Public Release

NOVEMBER 1986

AR-004-509

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A HYBRID HOT-WIRE DATA ACQUISITION SYSTEM

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SUMMARY

This report describes the philosophy, software, hardware and operation of a hybrid hot-wire signal processing system where unscaled mean and rms quantities are determined by analog circuits whose output voltages are passed to a digital computer for scaling and further processing. The circuits are orchestrated by a compact intelligent microprocessor based data acquisition system which can be supervised by any computer with an RS232C serial interface.

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ERRATA

Page 4. First equation should read :

$$E_{rms} = \frac{\sum\limits_{A=1}^{120} (E)^2}{120} - \left(\frac{\sum\limits_{A=1}^{120} E}{120} \right)^2$$

Page 22. Symbol $\boldsymbol{x}$ in final paragraph should be :

∂ (differential)

1. INTRODUCTION

Prior to the advent of the mini-computer in the mid-1960's turbulence signals from hot-wire anemometers were almost exclusively processed using analog techniques to produce the final quantities of interest. Laboratories wishing to process turbulence signals from hot wires typically possessed a combination of discrete component modules, such as chopper stabilised amplifiers, multipliers and integrators known as analog computers. Particular circuits could then be configured quickly as required by means of patch panels. Developments in digital computer technology since then have been such that microcomputers are available today that are more powerful than even the largest of machines available in the 1960s. Consequently most laboratories have discarded the analog methods in favour of purely digital processing techniques. While cheap and powerful microcomputers have removed much of the tedium involved with experimental signal processing the use of purely digital techniques has introduced some undesirable features into the methods used for measuring turbulence quantities.

The hot-wire anemometer is a frustrating instrument to use under the best of conditions. The output voltage consists of a relatively large dc component upon which are superimposed the voltage fluctuations of interest. The output voltage is a nonlinear function of both fluid velocity and fluid temperature. Temperature variations as small as 0.5 degrees centigrade are large enough to cause significant variations away from a voltage-velocity calibration. It is not unusual to experience extraneous drift in the calibration. Observation of hot-wire filaments under a microscope after several weeks of use in unfiltered environments usually will reveal black oxide coatings which are presumably formed by impact with minute droplets suspended in the air. For these reasons it is desirable to calibrate the instrument and take the measurements in as short a time as possible.

The nature of turbulence signals is such that a large number of discrete digital samples is required for adequate data convergence. In regions of high turbulence intensity the nonlinearity of the hot-wire response may require extra calculations to be performed at each data point which further slows up digital processing. Processing turbulence signals with analog circuits results in much faster data acquisition with an enormous reduction in data storage requirements compared to digital methods. Together with a reduction in cost there has been a dramatic increase in speed, accuracy and linearity and a reduction in the temperature sensitivity of amplifiers and computational devices in recent years.

This report describes the software, hardware and operation of a hybrid hot-wire signal processing system where unscaled mean and rms quantities are determined by analog circuits whose output voltages are passed to a digital computer for scaling and further processing. The circuits are orchestrated by a compact intelligent microprocessor based data acquisition system which can be supervised by any computer with an RS232C serial interface. The hybrid system is the result of continued development of the methods used by Watmuff, Witt and Joubert (1985). As well as increased analog circuit sophistication, the system has been automated and includes an informative screen display written for an industry standard VT100 terminal which shows the current operational status of the circuits in considerable detail. User friendly programs and informative screen displays have been written to aid and instruct the operator at every stage during the procedure involved. All commands to the system and all information requiring the operator's attention takes place via the single display terminal and keyboard. All the system hardware is contained in a portable rack that is easily transportable and the software is written in FORTRAN 77 under the RSX11-M operating system so that operation can be controlled by any DEC computer using RSX11-M with a spare RS232C serial printer port (e.g. the ARL Low Speed Wind Tunnel PDP-11/44 computer system).

2. HOT-WIRE ANEMOMETERS

Two dual channel constant-temperature hot-wire anemometers have been constructed by the author following the recommendations of Perry (1982). New improved low drift high frequency response JFET input (National LF411) operational amplifiers are used to provide the feedback gain in two stages with gains of 20 and 50 (i.e. a total gain 1000). The anemometer frequency response is inferred by injecting a 1 kHz square wave with a peak to peak amplitude of less than 200 millivolts via a front panel BNC connector. The frequency response is optimised to a value greater than 20 kHz by adjusting the second stage feedback amplifier offset voltage and the bridge inductance as provided for on the front panel. The reader is referred to Perry (1982) for detail of the theory and operation of the constant-temperature hot-wire anemometer which will not be described here.

The typical anemometer bridge output consists of a relatively large dc component upon which are superimposed the voltage perturbations of interest. Each of the new hot-wire channels contains an extra amplifier for subtraction of a dc 'bucking' voltage from the bridge output and four selectable gain levels of 1, 2, 5 and 10 for amplifying the resultant voltage fluctuations. The anemometers can be seen in figure 4.

3. HOT-WIRE CALIBRATION METHODS

The material contained in this section is intended as background information. Although much of it has been published in books and Journals some of the more recent developments remain in Ph.D. theses. It is included here since it is necessary to have at least a working knowledge of the concepts and jargon used in this report.

3.1. Static calibration methods

The hot-wire calibration procedure widely adopted around the world involves the use of a standard plot of the statically obtained quantities E_0^2 versus U^n, where E_0 is the mean output voltage, U is the mean velocity and n is an exponent, whose value depends on which heat-transfer law the user may favour. The usual form is given by:

$$E_0^2 = A + BU^n$$

The small perturbation sensitivity required for measurement of turbulence intensities is given by:

$$\partial E_0/\partial U = nBU^{n-1}/2E_0$$

which is determined after numerically fitting a line of best fit to the data points. This technique requires the operator to know the value of n which varies from one authority to another but is believed to be in the range 0.4-0.5, Collis and Williams (1959) have suggested that n=0.45 for straight wires with large length to diameter ratios. As a result of their work many operators have used n=0.45 rather than the original King's law value of n=0.5.

3.2. Dynamic calibration methods

An alternative procedure which avoids the need to presume the functional form of the heat-transfer law and which avoids the need to numerically differentiate the calibration curve is the dynamic calibration technique developed by Perry and Morrison (1971). This method involves shaking the hot wire in a uniform stream with an accurately known sinusoidal velocity perturbation at low frequency. Provided that the velocity perturbations are not too large, the small perturbation sensitivity of the wire can be determined directly with precision.

3.2.1. The dynamic calibrator

Since there is no need to presume a heat-transfer law for the hot-wire response the sensitivity of the amplified output of the anemometer can be used for the calibration. A dynamic calibrator has been constructed based on a machine used in the weaving industry for spinning self-twist yarn (Lavery and Henshaw 1972). Figures 1(a) to (d) are photographs of the calibrator. The machine operates on the principle of a Murray's cycloidal drive to provide simple harmonic motion when driven at constant speed. Balance weights are incorporated in the drive mechanism so that it and the associated mass being reciprocated (i.e. the sting carrying the hot-wire probe) are dynamically balanced. The variable speed dc motor drive input only has to cope with running friction and starting and stopping torques. The calibrator is capable of running at speeds in excess of 10 cycles/second but a speed of 2 cycles/second is adequate for most hot-wire work.

Ideally the calibrator should be attached to the tunnel in which the subsequent measurements are to be performed so that there is no need to switch off the anemometers or disconnect any leads. This arrangement also tends to minimise any temperature difference between calibration and run times. If it is necessary to calibrate and operate the anemometers in different tunnels where a temperature difference of more than 0.5°C is likely to occur the correction scheme devised by Abell (see Perry 1982) should be used. For the methods described here it is necessary to position the calibrator outside the working section such that the primary direction of sinusoidal oscillation is carefully aligned with the direction of the free-stream velocity in the working section as shown in figure 1(a). Two other directions of oscillation are provided for the calibration of crossed-wires. These directions are normal to and at 45 degrees to the free-stream velocity as shown in figures 1(b) and (c). It should be noted that it is necessary to transfer the probe back and forth between the calibrator and traverse stings. With crossed-wire probes considerable care is needed to align the probe with precisely the same orientation in both the shaker and traverse sting. Precision clamps are fitted to both stings and the shaker and traverse machinery have been carefully aligned during installation. Nevertheless the hot-wire voltages should be used to verify the alignment each time a probe transfer is performed.

3.2.2. Evaluation of the small perturbation sensitivity of a hot-wire signal

An optically encoded disc is fitted to the calibrator input shaft such that two optical transducers produce 120 pulses/cycle and 1 pulse/cycle of the calibrator motion. For the digital calibration techniques used at the University of Melbourne the 120 pulse/cycle signal is used to initiate analog-to-digital (A/D) conversions of the hot-wire voltages which are transferred into arrays in memory where they are added to those recorded during previous cycles. The single pulse/cycle signal is used to indicate the beginning of a new sampling cycle and to start and store an internal clock which is used to determine the shaking frequency. After a predetermined number of cycles (usually more than 50) the summed voltages corresponding to each of the 120 positions of the calibrator are averaged to remove the small random variations introduced by background turbulence. In other words the hot-wire signals are averaged on the basis of the phase (i.e. phase-averaged) of the calibrator motion. The rms value (E_{rms}) of the phase-averaged hot-wire voltages (E) are determined as:

$$E_{rms} = \frac{\sum_{A=1}^{120} (E)^2}{120} - \left[\frac{\sum_{A=1}^{120} E}{120}\right]^2$$

The rms shaker speed U_{rms} is calculated from the average shaking frequency (f) assuming pure sinusoidal motion, i.e.

$$U_{rms} \frac{A\ (2\pi f)}{\sqrt{2}} = \frac{A\omega}{\sqrt{2}}$$

where A = 38.1 mm is the amplitude of sinusoidal motion (i.e. the calibrator stroke) and ω is the frequency of oscillation in radians/sec. The ratio of these two quantities defines the hot-wire system sensitivity $\partial U/\partial E$ as,

$$\partial U/\partial E = U_{rms}/E_{rms}$$

In order to avoid the effects of hot-wire nonlinearity it is important to keep the magnitude of the rms velocity perturbation imposed by the oscillation less than around 10% of the tunnel free-stream velocity. The sensitivity determined this way is called the small perturbation sensitivity.

For the analog technique described here the rms hot-wire system voltage perturbation is obtained by electronic integration of the ac coupled and squared signal. It is necessary to filter the signal beforehand to remove the small perturbations introduced by background turbulence. The average shaking frequency is determined by counting the number of 120 pulse/cycle pulses over the integration time period which is accurately measured. Since the motion of the calibrator is not synchronised with the beginning and end of the integration time period the effect of a noninteger number of calibration cycles N needs to be considered. The mean square shaker speed is more correctly given by:

$$U_{rms} = \frac{(A\omega)^2}{T}\int_0^T \sin^2 \omega t\ dt = \frac{(A\omega)^2}{2} - \frac{(A\omega)^2}{2\omega T}\int_0^{\omega T} \cos 2\omega t\ d(\omega t)$$

For $\omega T = 2N\pi$, the integral of the cosine term is zero and the actual and calculated calibrator speeds are in exact agreement. The maximum error in mean square shaker speed due to the assumption of an integer number of calibration cycles ϵ_{max}, is given by,

$$\epsilon_{max} = \frac{1}{2N(2\pi)}\int_{-\pi/4}^{\pi/4} \cos 2\omega t\ d(\omega t) = \frac{1}{4N\pi}$$

Thus if the integration time period is greater than 16 times the shaking period for example, the maximum error in U_{rms} is less than 0.25% and can be neglected.

3.2.3. The small perturbation sensitivity of a crossed-wire probe.

If an inclined wire is oscillated sinusoidally in a steady flow along the flow direction the streamwise sensitivity $\partial E_0/\partial U$ may be determined. If it is oscillated across the stream the cross-stream sensitivity $\partial E_0/\partial V$ may be evaluated. The Reynolds stress sensitivity of an inclined wire may be found by oscillating the wire in a path inclined to the mean flow direction since the wire experiences simultaneous streamwise (u) and cross-stream (v) velocity fluctuations. The small perturbation sensitivity of a wire inclined at an angle α to the free-stream is given by,

$$e_{01} = S_1 u + S_2 v \tag{1}$$

where, $S_1 = \partial E_0/\partial U$ and $S_2 = \partial E_0/\partial V$

If the wire angle is changed to $-\alpha$ by rotating it by 180 degrees about an axis parallel to the mean flow the output voltage perturbation is,

$$e_{02} = S_1 u - S_2 v \tag{2}$$

A combination of mean-square values of equations (1) and (2) gives,

$$\overline{uv} = (\overline{e_{01}^2} - \overline{e_{02}^2}) / 4S_1S_2$$

where overscores denote temporal mean values.

It can be shown that a single inclined wire which can be <u>rotated</u> by 180 degrees <u>is</u> suitable for measuring Reynolds stress only. To measure $\overline{u^2}$ and $\overline{v^2}$ as well as $\overline{uv}$ it is necessary to use a crossed-wire probe. A crossed-wire probe can be thought of as an inclined wire with a counterpart '180 degrees rotated' inclined wire placed closely beside it with the angle α being approximately 45 degrees.

The principles of dynamic calibration of crossed-wire probes were initially developed by Morrison, Perry and Samuel (1972) and have been recently extended by Watmuff (1979), Perry (1982) and Tan (1983). Only a brief outline will be given here. The approach is based on a Taylor series expansion of the hot-wire response equations to obtain a power series of arbitrary order relating the voltage output to velocity. The method does not require any measurements of wire angles nor are precise functional forms assumed for the heat-transfer laws. However these heat-transfer laws are useful for producing an analysis which acts as a guide for illustrating the procedures involved. Similar end results can be obtained using more complex relationships for the heat transfer. For simplicity the unmodified cosine cooling law and Kings law with an exponent of 0.5 will be used in the following linearised analysis. Consider two crossed-wires located in a two-dimensional flow field as shown in figure 2. Using the notation shown in the figure the outputs of the two anemometer channels are ,

$$E_{01}^2 = A_1 + B_1 \left[(U_R + u)\cos\phi_1 + v\sin\phi_1 \right] \tag{3(a)}$$

and

$$E_{02}^2 = A_2 + B_2 \left[(U_R + u)\cos\phi_2 - v\sin\phi_2 \right] \tag{3(b)}$$

where U_R is an arbitrary streamwise reference velocity about which u and v are perturbations. Taking the square root of both terms, it can be shown that,

$$E_{o1} = E_{R1} + \frac{B_1\sqrt{\cos\phi_1}}{4E_{R1}\sqrt{U_R}}\,u + \frac{B_1\sqrt{\cos\phi_1}}{4E_{R1}\sqrt{U_R}}\,v\tan\phi_1, \text{ where } E_{R1} = [A_1 + B_1\sqrt{U_R}\cos\phi_1]^{1/2} \qquad 4(a)$$

and,

$$E_{o2} = E_{R2} + \frac{B_2\cos\phi_2}{4E_{R2}\sqrt{U_R}}\,u - \frac{B_2\sqrt{\cos\phi_2}}{4E_{R2}\sqrt{U_R}}\,v\tan\phi_2, \text{ where } E_{R2} = [A_2 + B_2\sqrt{U_R}\cos\phi_2]^{1/2} \qquad 4(b)$$

It can be seen that u and v may be found from the sum and the difference of E_{o1} and E_{o1} respectively. The 'bucking' voltages E_{B1} and E_{B2} which are part of each anemometer channel are adjusted so that when the signals are amplified by K_1 and K_2 the output voltages are given by,

$$E_1 = K_1(E_{R1} - E_{B1}) + \frac{K_1B_1\sqrt{\cos\phi_1}}{4E_{R1}\sqrt{U_R}}\,u + \frac{K_1B_1\sqrt{\cos\phi_1}}{4E_{R1}\sqrt{U_R}}\,v\tan\phi_1 \qquad 5(a)$$

$$E_2 = K(E_{R2} - E_{B2}) + \frac{K_2B_2\sqrt{\cos\phi_2}}{4E_{R2}\sqrt{U_R}}\,u + \frac{K_2B_2\sqrt{\cos\phi_2}}{4E_{R1}\sqrt{U_R}}\,v\tan\phi_2 \qquad 5(b)$$

By using the potentiometers (as shown in the circuit diagram in figure 3) to attenuate the voltage E_2 by the factors K_2' and K_2'' which are given by,

$$K_2' = \frac{K_1' \quad K_1B_1E_{R2}\sqrt{\cos\phi_1}\tan\phi_1}{K_2 \quad B_2 \quad E_{R1}\sqrt{\cos\phi_2}\tan\phi_2}, \qquad K_2'' = \frac{K_1' \quad K_1B_1E_{R2}\sqrt{\cos\phi_1}}{K_2 \quad B_2 \quad E_{R1}\sqrt{\cos\phi_2}}$$

and separately summing and subtracting the resultant signals from E_1, then E_u and E_v are given by,

$$E_u = \frac{K_1B_1\sqrt{\cos\phi_1}}{4E_{R1}\sqrt{U_R}}\left(\frac{\tan\phi_1 + \tan\phi_2}{\tan\phi_2}\right)u \qquad 6(a)$$

$$E_v = \frac{K_1B_1\sqrt{\cos\phi_1}}{4E_{R1}\sqrt{U}}(\tan\phi_1 + \tan\phi_2)\,v \qquad 6(b)$$

Hence to the linearised approximation E_u and E_v are proportional to u and v alone respectively. The potentiometer K_1' is included in the matching circuit (fig. 3) for reducing the magnitude of the sensitivities. It must be set and held fixed before and after the matching procedure.

The procedure of adjusting the analog (matching) circuit while the wires are being oscillated is known as 'matching the wires'. In practice matching should be performed at a tunnel velocity midway between the two extremes to be experienced by the wires. As the probe is oscillated across the free stream the potentiometer K_2' is adjusted to make the voltage E_u optimally independent of the velocity fluctuations by observing the signal on an oscilloscope. Similarly potentiometer K_2'' is adjusted to minimise the cross contamination of the voltage E_v while the probe is oscillated in the steamwise direction. Perfect matching is not attainable but the ratio of wire sensitivities to the two velocity components (or the rejection ratio) is typically 50:1. Experience has shown that for physically similar wires the matching remains valid over a wide range of velocities. Validity tests should be performed during the matching operation by varying the tunnel velocity over the anticipated range of velocities after completion of each of the streamwise and cross-stream shaking operations. The same analog circuit and matching procedure is used before both analog and digital calibrations.

3.2.4. Full nonlinear calibrations derived from small perturbation sensitivities.

The small perturbation sensitivity of a hot-wire voltage can only be used for the evaluation of turbulence quantities if the rms turbulence intensity is less than around 10% of the mean velocity. The advantages of using the small perturbation sensitivity method is that voltages do not have to be converted to velocities before evaluation of turbulence quantities is performed. Instead only voltages need to be processed from which the rms velocities are calculated afterwards. For rms turbulence intensities above 10% the effects of hot-wire nonlinearity should be taken into account. It is not possible to provide a definitive value for upper limit on the turbulence intensity without knowing the probability density function (pdf) of the turbulence signals. However the evidence suggests that an upper limit of 10% will be safe for most turbulent flows. Provided this condition is satisfied then the small perturbation sensitivity has to be known at each experimental operating point. During a boundary layer traverse, for example, a different mean velocity (and hence a different operating point) is experienced at each of typically 40 or 50 traverse positions from the wall. It is obviously much too time consuming to evaluate the small perturbation sensitivity at each of these operating points. Instead the small perturbation sensitivities can be determined from a curve fit applied to a lesser number (.e.g. usually 6 to 10) of accurately determined calibration values spaced at approximately equal intervals over the range of velocities to be experienced. A second-order least squares polynomial of best fit is adequate for most applications.

The technique of curve fitting the small perturbation sensitivities as a function of the mean hot-wire system voltage can be extended to provide a full non-linear calibration for the evaluation of turbulence quantities in regions where the rms turbulence intensities are greater than 10%. The numerical scheme for determining the full nonlinear calibration for both the normal and cross-wires is independent of whether the small perturbation sensitivities have been determined by analog or digital methods. For a normal hot-wire the polynomial fit to the system sensitivities can be analytically integrated to give the full nonlinear cubic calibration polynomial.

$$U = A + BE_u + CE_u^2 + DE_u^3$$

$$\partial U/\partial E_u = B + (2C)E_u + (3D)E_u^2$$

The constants B, C, and D are determined from the curve fit to the sensitivities while the unknown constant of integration A is determined as the value that gives the least square deviation of the cubic polynomial from the free-stream velocities which are determined by a Pitot-static tube at each calibration operating point. It is important to realise that the cubic calibration polynomial is derived from the small perturbation sensitivities. While the hot-wire output may drift with time causing significant errors in mean velocities that may be inferred from the calibration, the effect on the velocity perturbations is of second-order and negligible. It is highly recommended that if a normal wire is to be used for mean velocity measurements, then a calibration should be performed before and after each experimental run and the results rejected if the two calibrations differ by more than an acceptable percentage.

The following technique of determining a full nonlinear calibration for crossed-wires was developed by the author in collaboration with Professor A.E. Perry for the purpose of evaluating the phase-averaged velocity vector fields behind three-dimensional bodies (Perry and Watmuff 1981). Since then contributions have been made by Perry (1982), Tan (1983) and Steiner (1984). The small perturbation sensitivity of the matching circuit voltage E_u can be determined at a number of representative velocities (as for a normal wire) by shaking the probe in the streamwise direction while the small perturbation sensitivity of E_v can be determined by imposing cross-stream velocity perturbations. The small perturbation sensitivities to streamwise and cross-stream velocity fluctuations and to Reynolds stress fluctuations can then be determined directly and simultaneously. All calibration data reduction code has been written assuming this 45 degree shaking direction. Once again it is important to keep the rms velocity perturbation less than 10% of the tunnel free-stream velocity to avoid the the effects of hot-wire system nonlinearity. The unperturbed free-stream velocity is measured at each operating point by a Pitot-static tube.

The linearised matching theory described earlier can be extended to arbitrary order series expansions (see Tan 1983). Satisfactory calibration curve fits can be realised by third-order polynomials in practice. The best results for the turbulence quantities are achieved by fitting the derivatives of the calibration curve rather than fitting the static data points. Therefore second-order least squares polynomial fits are applied to the small perturbation sensitivities $\partial U/\partial E_u$ and $\partial V/\partial E_v$ as functions of the mean voltage E_u at each calibration point, i.e.

$$\partial U/\partial E_u = B + (2C)E_u + (3D)_u^2$$

and

$$\partial V/\partial E_v = R + SE_u + TE_u^2$$

Note that no attempt is made to include the effects of nonlinearity of E_v for large cross-stream velocity fluctuations. Theoretical modelling and experiments performed by Tan (1983) and Steiner (1984) have shown that the response is linear with respect to E_v provided that the wedge angle of the velocity vector remains within ±30 degrees for 90 degree crossed-wires. For wedge angles greater than about ±30 degrees the behaviour predicted by the computer modelling indicates that the nonlinearity needs to be taken into account. If the wedge angle exceeds either of the inclined wire angles then the velocity is indeterminate because of directional ambiguity. In practice small vortices are shed from the stubs and prongs long before this limit is reached. These vortices interfere with the flow over the filaments causing high frequency voltage oscillations even though the wires are in a steady flow.

The expressions for the sensitivities can be analytically integrated to give the full nonlinear calibration polynomials as,

$$U = A + BE_u + CE_u^2 + DE_u^3$$

$$V = \pm P \pm QE_u + RE_v + SE_u E_v + TE_u^2 E_v$$

The unknown constant of integration A is determined in the same way as for the normal wire. The unknown integration constants P and Q are evaluated in a similar manner by assuming that the mean cross-stream velocity is zero during the shaking procedure. The values of P and Q are small in practice for physically matched wires. Tan (1983) has shown that the second- and third-order terms that are not included in the polynomial for V are very small and can be neglected. He also showed that these relationships remain quite accurate even with imperfectly matched wires.

4. COMPARISON OF ANALOG AND DIGITAL METHODS FOR PROCESSING HOT-WIRE SIGNALS

Because of the 'spikey' nature of turbulence signals a large number of discrete digital samples is needed for adequate data convergence. Continuous data sampling at a rate of 10 kHz over a period of 30 seconds is not an unusual requirement. If the rms velocity is less than around 10% of the mean velocity then the rms voltage need only be measured. The rms velocity can be calculated afterwards from the small perturbation sensitivity. Even when the fluctuations are small enough for the small perturbation sensitivity to be used, processing hot-wire signals with digital methods is time consuming because of the large quantity of data that is involved. In regions where the turbulence intensity is larger than 10% (e.g. close to the wall in a turbulent boundary layer) the assumption of hot-wire linearity breaks down and rms velocities calculated with the small perturbation sensitivity method will be in error. As mentioned previously it is difficult to provide an estimate of the error without knowing the pdf of the signals. Henbest (1983) performed a series of tests in a smooth-walled pipe where time sequences of digitally sampled normal hot-wire voltages were recorded on magnetic tape and subsequently processed using two methods. In the first method the rms voltage was calculated and converted to the rms velocity using the small perturbation sensitivity. In the second method each voltage was converted to a velocity using the full nonlinear calibration to calculate the rms velocity directly. Although the results from the two methods showed substantial agreement in the outer region of the flow discrepancies of up to 4% were observed when the probe was placed close to the wall. While providing greater accuracy, the use of the full nonlinear calibration requires further calculations to be performed for each data point which further slows up digital signal processing. Laboratory microcomputers are not capable of repeatedly sampling crossed-wire voltages, converting the measurements to velocities through the full nonlinear calibration polynomials and performing the necessary arithmetic to produce the final results of interest at a rate that is sufficiently high for turbulence signals. Benchmark tests for crossed-wire signal processing indicate that the current generation of 16-bit processors can cope with less than 500 samples/sec. Before large quantities of memory became available a number of small ensemble time records were stored and then processed. With the recent availability of large quantities of cheap memory (e.g. 4 megabytes on a single DEC Q-bus module) continuous time sequences of raw data can now be gathered. However subsequent on-line processing causes considerable time delay before the next data record can be obtained. One way of avoiding the excessive time delays caused by the calibration inversion and rms velocity calculations is to store the sampled data on a peripheral device such as a disc or tape. However the data still has to be processed at a later date often without the knowledge of whether enough samples have been obtained for adequate data convergence. With this method the collection of a large set of data files requiring up to 30 megabytes of disc or tape storage per boundary layer profile for example is not uncommon. Up to three hours of post-experimental data processing is still required to produce the results of interest.

Processing turbulence signals with analog circuits results in much faster data acquisition and an enormous reduction in data storage requirements compared to digital methods. The final results of interest are available for inspection almost immediately upon completion of the measurement period. Together with a reduction in cost there has been a considerable increase in the frequency response, accuracy and linearity combined with a reduction in temperature sensitivity of amplifiers and computational circuits in recent years. Calibration of analog circuits can be carried out with precision by applying accurately known inputs and measuring the corresponding outputs. Compensation for small offset voltage variations can be made by frequent measurement so that device nonlinearity becomes the dominant source of error. Devices are readily available that are linear to better than 0.1% of their full-scale outputs.

Because of the speed and accuracy of processing turbulence signals with analog techniques a hybrid hot-wire signal processing system has been devised where unscaled temporal mean and rms voltages are determined by hard-wired analog computer circuits. A digital computer is used to measure and scale the voltages and perform the circuit calibration and data reduction calculations so that the results are available almost instantly after completion of the measurement process. The computer is also used to monitor and control the circuits so that the procedures may be automated. The circuits are the result of continued development of those used by Watmuff, Witt and Joubert (1985). The methods used by Watmuff et al were thoroughly tested against the corresponding digital techniques developed at the University of Melbourne (utilising a PDP 11/10 minicomputer fitted with a Laboratory Peripheral System) by applying the same calibration and experimental crossed-wire signals to both systems. The results were always within a few percent of each other. The hard-wired analog computer system was not only an order of magnitude faster but also gave more repeatable results. As well as increased circuit sophistication, the system developed here has now been automated and includes a continually updated informative screen display which dynamically shows important voltage values and the operational status of the circuits in considerable detail. All commands to the system and all information requiring the user's attention take place via a single display terminal and keyboard.

5. COMPONENTS OF THE HYBRID DATA ACQUISITION SYSTEM

5.1 Digital Computer

The microcomputer used in the hybrid hot-wire signal processing system is an MDB Systems Inc. Micro/11 which has a 22-bit Digital Equipment Corporation (DEC) compatible Q-bus backplane that can accommodate a total of eight quad or sixteen dual size circuit board modules. A DEC PDP 11/23-PLUS Central Processing Unit (CPU) is provided with a Memory Management Unit that supports 22-bit addressing (i.e. up to four megabytes of memory can be installed). The quad-height CPU module includes two asynchronous RS232C serial interfaces, a bootstrap loader, diagnostics and a programmable line frequency clock. The optional KEF11-AA floating point chip is also included.

A 5¼ inch Winchester fixed disc is supplied which emulates two 10 megabyte DEC RL02 drives. A dual-height disk controller which is fully transparent to the DEC RL02 driver software provides the Direct Memory Access (DMA) interface. Two 8 inch Floppy Disks with a total capacity of 1 megabyte provide the only removable media i.e. the Floppy Disks are the only means of archiving data and source programs. The Floppy Disk drive is RX02 media compatible allowing data and program exchange between most other DEC systems. A separate dual-height Floppy Disk controller which is fully transparent to the DEC RXV21 driver software provides the DMA interface.

In it's current configuration the system is equipped with an MDB supplied dual-height 256 kilobyte parity memory module. This will be shortly replaced by a

Fourth Generation Systems quad-height four megabyte memory module. These memory modules are fully compatible with the DEC block mode DMA.

A dual-height Webster Electronics 8-port asynchronous serial line multiplexer has been installed providing access to the system for devices using standard RS232C protocol. The multiplexer is functionally equivalent to the DEC DZV11 but offers twice as many serial lines while only using half the backplane space. The module features software selectable baud rate, character length, number of stop bits and parity checking for each serial line and provides access to the newer faster transmission rates of 19200 and 32400 bits/sec. Together with the (non-programmable) lines on the CPU board the system has a total of ten serial interface lines. A C-ITOH 101-e (DEC VT100 compatible) video terminal and a Facit 4512B 150 character/sec dot matrix line printer are connected to the CPU board serial interfaces. One of the multiplexer ports is used to support a microprocessor-based data acquisition system and another is used for a small HP7470 digital plotter. The remaining serial ports provide for future expansion and are presently used for general purpose video terminals which are connected from time to time to take advantage of the multi-user operating system.

5.2 Computer Operating System

The operating system chosen to support the hybrid hot-wire data acquisition system software is Digital Equipment Corporation's (DEC) RSX11-M which is a multiuser multitasking system that has been optimised for real-time operation. Concurrent execution of several tasks residing in memory (multitasking) is possible since while one task is waiting for completion of an input/output operation (for example) another task can have control of the central processing unit (CPU). Task execution is event driven using a priority ordered queue under control of the operating system executive. Task priorities maybe set in the range of 1 to 250. A task retains control of the CPU until it is interrupted by a task of higher priority or until it is unable to continue while waiting for completion on an input/output operation. A part of the executive called the scheduler monitors the system at regular intervals (every 1/50th of a second) to ensure that tasks of equal priority are allocated an equitable share of system resources i.e. no one task can dominate the CPU until they are finished.

The executive also provides the software interfaces between user programs and the system hardware i.e. memory allocation, device drivers, file management, system utilities and programmed system services (called executive directives) are all provided by the executive. The executive directives offer powerful features to the real-time programmer in the form of FORTRAN callable subroutines. Directives are available which allow one task to start another task, to repetitively run another task at predetermined intervals or to stop the execution of another task. A task can be given the ability to dynamically alter the priority of another task or of itself. Memory can be partitioned into regions which are globally common to a group of tasks in much the same way as a COMMON statement in FORTRAN allows subroutines of a main program to share memory. Communication between tasks can be synchronised by means of event flags which may be set by the executive to indicate the completion of an input/output operation for example or which may be set or cleared by a task to signal to other tasks the completion of an operation.

Of special significance to the hot-wire signal processing system are the directives available for input/output communications. These directives allow a task to access devices at the driver interface level providing increased speed and the ability to perform operations not normally available to the FORTRAN programmer. Extensive use has been made of system traps or software interupts which are means by which a task can monitor and respond to events. Synchronous System Traps (SSTs) always occur at the same point in a program and are taken care of by the executive with no intervention needed by the user. One other other hand Asynchronous System Traps (ASTs) detect and inform the task of events over which the task has no a priori knowledge of the precise time at which the event will occur. ASTs are dealt with by means of a short user-written subroutine (usually in macro assembler) to which control is transferred when the AST occurs.

The author is indebted to Dr. B.D. Fairlie for his willing advice and help with the operating system on numerous occasions and for the provision of an AST for monitoring unsolicited input from terminals. The unsolicited input AST sets an event flag and any other characters typed on the terminal are stored in a buffer. Meanwhile the task can continue to execute while occasionally checking the status of the event flag using the operating system directive READEF. When the event flag is set the task can retrieve the characters from the buffer, carry out any coded processing indicated by the characters and then continue to execute from where it was interrupted. The unsolicited input AST is especially useful when a task is performing repetitive experimental operations. For example if an undesirable situation develops while an experiment is being monitored the operator need only type a carriage return to halt (or interrupt) program execution. Typically a menu can be presented to the operator from which the best course of action can be selected for remedying the situation.

5.3 Integrator Circuit

A single printed circuit board (PCB) containing the necessary hard-wired analog computer circuitry has been designed and built for determining temporal mean velocities and root-mean-square (rms) and Reynolds stress turbulence quantities from signals derived from hot-wire probes. The PCB contains a total of eight integrators. One of the integrators is dedicated for setting the adjustable integration time period. Four general purpose integrators (labelled DC) are provided for the determination of true mean voltages. The first two DC integrators are usually designated for the hot-wire voltages while the other two can be used for miscellaneous purposes such as evaluation of mean outputs of pressure transducers for example.. The input to each integrator is attenuated via a front panel potentiometer in order to avoid integrator overloads as the range of input voltages and the integration time period are varied for different applications. The remaining three integrators (labelled RMS) are dedicated for the determination of rms and Reynolds' stress turbulence quantities.

Each of the two inputs to the RMS integrator circuitry enters a two stage amplifier system. The first stage amplifiers can be ac or dc coupled via a toggle switch on the front panel. The ac coupled roll-up frequency is around 0.5 Hz. It is essential that the phase response of each ac coupled amplifier be identical for accurate determination of the Reynolds stress so the phase response has been carefully tuned using the Lissajous figure technique. Two second stage variable gain dc amplifiers are provided for each of the first stages. One is for the respective squarer and the other is for the multiplier inputs. All four second stage amplifiers have four gain levels of 1, 2, 5 and 10 which are independently selectable via front panel rotary switches. Front panel sockets are provided for monitoring the fluctuating outputs of the second stage amplifiers, squarers and multiplier on an oscilloscope. Potentiometers are also mounted on the front panel between the multiplier/squarers and their respective integrators for the purpose described earlier. The front panel of the integrator circuit can be seen in figure 4.

The circuit is controlled either remotely by digital inputs or manually by a front panel rotary switch with three positions labelled reset, hold and integrate. Local (manual) or remote operation is selected by a front panel toggle switch. Due to the arrangement of PCB mounted analog switches only two control lines are needed i.e. reset and integrate. When the rest control line is activated the analog switches are toggled to connect a 10KΩ dumping resistor in parallel with the feedback capacitor of each integrator causing the integrators to discharge. When the reset line is deactivated (e.g. the switch is thrown to the hold or integrate positions) the dumping resistors are disconnected. Integration is determined by the timing integrator whose output is used to fire a Schmit Trigger with fixed threshold voltages. The integration time period is varied by adjusting a precision voltage fed front panel potentiometer which provides the input voltage. Each input to the integrators is connected via one pole of a two pole analog switch whose opposite pole is connected to ground. The complementary output of the Schmitt Trigger is used to toggle these analog switches via another analog switch which is toggled by the integrate control line. This arrangement ensures that all integrator inputs (including the timing integrator) are grounded when the integrated control line is deactivated (e.g. when the rotary switch is in the reset or hold positions). As the output of the timing integrator rises it fires the Schmitt Trigger causing the complemtary output to go low thus connecting the integrator inputs to ground and terminating the integration process. Two status lines are used to indicate circuit operation and drive lights on the front panel labelled 'integrating' and 'finished'. Measurements with a Hewlett Packard Digital Timer indicates that the integration time period (typically 30 sec) is repeatable to within ±100 microseconds.

5.4 Microprocessor based control and data acquisition system

The PDP 11/23-PLUS microprocessor is used to control and monitor the operation of the integrators and to measure and process circuit voltages via a data logger with the brand name of Dataporte. The Dataporte is a compact low-cost intelligent microprocessor-based data acquisition system which can be supervised by any computer with an RS232C serial interface. Amongst the features of the Dataporte are analog, digital, event and counter inputs, data storage and averaging, analog and digital outputs, temperature measurement, and a real time clock. Figure 5 is a photograph of the Dataporte.

Up to 46 single-ended analog inputs can be connected so that it is possible to monitor the input and output of every major device on the integrator and lineariser circuit boards. The Dataporte is authoranging over the three voltage ranges ±25mv, ±250mv, ±2.5v. For devices whose voltages are likely to exceed ±2.5v (e.g. the seven integrators) it is necessary to attenuate the Dataporte inputs with simple two resistor voltage dividers. Each attenuation factor has been determined by applying an approximately equally spaced sequence of voltages to the attenuators over the range to be experienced. A least square line of best fit has been applied to the attenuated voltages measured by the Dataporte and to the voltages measured by a Hewlett Packard digital voltmeter (DVM). The maximum deviation from the lines of best fit from the DVM readings was found to be within ±0.5mv over a range of ±10v for all attenuators.

The Dataporte has 8 digital input/output channels which are used to control and monitor the integrator circuit. Two digital outputs are used to start and to reset the integrators while two digital inputs are used to detect the 'integrating' and 'finished' status lines. Four digital outputs are used to control the 16-channel analog multiplexers of the autocal circuit which has been designed to supply the integrator inputs during automated circuit calibration. A fifth digital output is used to toggle a set of analog switches which connect either the front panel inputs or the autocal circuit voltages to the circuit. The last digital output is used to toggle a set of two pole analog switches that disconnect all inputs and ground the devices on the circuit board. This is required during the automated circuit calibration and is also used before each integration cycle to monitor and correct for offset voltage drift. Also included in a high speed (2 MHz) counter with a gated input connected to the integrating status line. The high speed counter is fed by an external precision crystal controlled oscillator so that the integration time period can be determined with precision. Each of the 8 digital inputs may act as low speed counters which are used in the hot-wire calibration procedure.

All communications to and from the Dataporte are in standard ASCII format. The Dataporte is controlled by issuing English-like commands which maybe abbreviated to single upper case letters e.g. 'R37V' which causes repeated sampling and return transmission of the floating point value of the voltage connected to analog input channel 37 without any further intervention. Several command sequences may be joined together to form a powerful command line of up to 126 characters. Frequently measured groups of voltages (e.g. the seven integrator outputs) have been connected to the Dataporte with sequential channel numbers so that coded scan commands can be issued which speeds up the measurement process. Switches maybe appended to commands to allow the Dataporte configuration to be changed to suit particular requirements and conditions. Other commands are available for altering the contents of particular internal program memory locations or to directly access some of the hardware features. For example the format of returned data may be tailored for convenient processing with FORTRAN subroutines by selecting appropriate data delimiter and end-of-line characters.

The maximum serial data rate is 4800 bits/sec and the maximum sampling rate is around 20 samples/sec so that the Dataporte is only suitable for applications requiring low sampling data rates. Nevertheless the response times are such that after a FORTRAN WRITE statement has sent a command to be transmitted on the printer port the following FORTRAN READ is too slow to be able to pick up the complete string of incoming characters. Extensive use has been made of FORTRAN callable RSX11-M operating system directives (e.g. SUBROUTINE WTQIO). The availability of 46 analog input and 8 digital input/output channels, the high accuracy and autoranging capability, the simplicity of programming and the portability and convenience of utilising an ordinary serial interface combine to make the Dataporte a versatile component for use with the hybrid hot-wire signal processing system where high data sampling rates are not required.

5.5 Integrator autocalibration reference voltage source

After the integrator circuit potentiometers have been adjusted and fixed to accommodate the range of anticipated inputs for a given integration time period, the DC and RMS integrators are calibrated by applying accurately known dc input voltages and accurately measuring the integrator outputs at the end of the integration time period. The first stage input amplifiers to the RMS integrators must be dc coupled for the calibration which must be performed for every combination of each of the four gain levels of the four second stage amplifiers that will be used for subsequent measurements. Although there are 256 possible combinations of gain levels this number of combinations will never be used in practice. For most measurements the gain levels of all four amplifiers are best kept the same. Occasionally the need may arise to use different gain settings depending on the nature of the signals but it is recommended that the same gain levels be used within each group of second stage amplifiers.

Three adjustable dc reference voltages have been provided on the integrator and lineariser circuit boards for calibration purposes. These voltages must be adjusted after each calibration integration cycle as the combination of gain levels are selected. Since all the integrators operate in parallel the opportunity exists to calibrate the RMS and DC integrators simultaneously. When the circuits were conceived it was envisaged that two of the dc reference voltages could be used for inputs to the RMS integrators leaving the other reference for an input to a DC integrator. Since the DC integrators can be satisfactorily calibrated by a single input/output measurement, the reference voltage can be suitably adjusted and applied to each DC integrator in succession as the two RMS integrator inputs are adjusted to accommodate the four gain settings.

A complete set of hot-wire measurements is obtained over a period of time typically measured in weeks. It is advisable to perform a circuit calibration once a day before performing the hot-wire calibrations and experimental runs. The procedure of adjusting the input reference voltages through the same sequences each day is inconvenient and wasteful of time. Therefore the 'autocal' circuit was devised for the purpose of supplying a complete set of reference voltages for simultaneous calibration of both the DC and RMS integrators. The autocal circuit parameters need only be adjusted once to suit a particular setting of the integrator circuit parameters. The front panel of the autocal circuit can be seen in figure 6.

The integrator reference voltages derived from the autocal circuit are selected under the control of the microprocessor software via four of the Dataporte digital outputs. The two reference voltages for the RMS integrators are derived from a master reference voltage which is set via the front panel. A knob labelled gain has four positions corresponding to the four gain levels of the RMS integrator front panel. The label is in fact a misnomer since the order of the gain levels are inverted with respect to those on the integrator front panel. This has been incorporated deliberately so that when the gain level positions on the autocal and integrator front panels correspond, the multiplier/squarer outputs are around the same values. Only two of the Dataporte digital outputs are used to control the RMS integrator reference voltage sources. For each gain level the digital outputs toggle analog switches so that a set of four reference voltages of different sign and magnitude are supplied for automated four quadrant calibration of the squarers and multiplier. Although a minimal circuit calibration requires only one input voltage and a single integration per circuit gain level the provision of a set of four reference voltages offers more thorough checking of the system accuracy. For the DC integrators a minimal calibration requires only one integration cycle. However since up to 16 integration cycles can be used for the RMS integrators advantage can be made of the simultaneous operation of the DC integrators by applying up to 16 evenly spaced dc reference voltages over the range to be experienced. The calibration reference voltages for the DC integrators are obtained by multiplexing voltages from a ladder network of resistors under the control of the Dataporte. The maximum and minimum voltages applied to the resistor network are adjusted on the front panel. The application of up to 16 calibration voltages to the DC integrators consumes no extra time and provides a thorough check of the system linearity.

5.6 Analog hot-wire signal lineariser circuit

In order to eliminate one source of error in regions of high turbulence intensity an analog hot-wire signal lineariser has been constructed in the form of a single printed circuit board containing three multipliers, a squarer and associated amplifiers. For convenience the analog crossed-wire matching circuit is also included on the board. The lineariser is in the form of two polynomial function generators designed to follow the nonlinear cubic calibration polynomials for E_u and E_v. Since the lineariser components are designed to operate over a voltage range of ±10 volts a linear scaling factor and offset voltage may need to be used. This will not alter the primary function of the lineariser which is to provide an output voltage that is linearly related to the velocities. For the U lineariser channel a constant dc voltage and the signals E_u, E_u, and E_u are connected to the front panel potentiometers (which supply the inputs to the final summing amplifier) via front panel switches so that each of the amplifier inputs can be grounded. The value of each polynomial coefficient is set by adjusting the respective potentiometer while all the other amplifier inputs are grounded. The same technique is used to set the V channel lineariser. The front panel of the lineariser and matching circuits can be seen in figure 4.

The typical 1960's analog computer multiplier used switching diodes to provide a piecewise approximation to their function. This type of multiplier is unsuitable for use in the lineariser circuits since even small discontinuities in the derivatives of the input/output relationship would have a detrimental effect on the calibration scheme. In order to test the circuit the small perturbation sensitivity calibration method was applied to the lineariser channels which were set to follow a set of typical crossed-wire calibration polynomials. The simulated calibration was achieved by applying to the lineariser input a small amplitude sine wave which was superimposed on a variable dc voltage. Ten different dc voltage values were used. the full nonlinear calibration polynomials derived from the small perturbation sensitivities save calculated values within 0.5% of the cubic polynomial of best fit which was found by applying the dc voltages only.

It is not recommended that the output of the lineariser be used for measurements after being set to follow the calibration polynomials. Experience has shown that small errors may accumulate throughout the system to produce results that depend on the operator's accuracy and patience during the set up procedure. Instead it is good practice to repeat the calibration but with the whole system, including the lineariser, before taking the measurements. As well as providing feedback on system linearity the effects of accumulated errors are made negligible. It may seem an excessive requirement to perform two calibrations in succession i.e. one for the matched crossed-wire voltages to that the lineariser can be set, and another afterwards for the lineariser outputs. However in practice this dual calibration requirement need only be performed occasionally for a given pair of crossed-wires. Although a series of calibrations may differ slightly from day to day due to temperature variations for example, it has been found that the matching circuit potentiometers do not need to be altered since the lineariser sensitivities remain approximately constant. Therefore the lineariser output alone may be calibrated for future measurements without any further adjustments. If the small perturbation sensitivities of the lineariser output begin to vary by more than about 10% after some period of time then it is recommended that the lineariser be set with the new coefficients from a fresh calibration of the matched crossed-wire voltages.

6. HYBRID DATA ACQUISITION SYSTEM SOFTWARE

6.1 Device controlling/measuring subprogram modules.

The hybrid hot-wire signal processing system device controlling software has been developed into a hierarchical framework of subprogram modules in the form of a relocatable object code library. The simplicity and clarity provided by a modular approach makes existing programs easy to understand and modify. Running programs are compact since the same module may be used many times with different parameters to perform a variety of functions. The program developer is relieved of the concern with the particulars of which circuit board devices are connected to which Dataporte channels so that concentration can be aimed at developing the task at hand. For new applications complete programs can be developed quickly by connecting sequences of easily remembered modules together.

At the lowest level the software consists of a number of self-contained subroutines each of which performs a specific primary function. All commands issued to the Dataporte are such that a single reply is required before proceeding. At no stage are commands issued that may lead to a delayed or multiple reply even though this means that certain powerful features of the Dataporte are not utilised e.g. sample upon a digital event or repeated sampling of analog inputs. The primary functions of the lowest level subprograms are apparent from their titles. Amongst the lowest level integrator control/status subprograms are :

GNDINP - activates the analog switches that disconnect all inputs to the integrators and ground them on the PCB.

INTGRT - activates the integrate control line.

RESET - deactivates the integrate control line and activates the reset control line for approximately one second during which time the high speed counter is reset for the next integration cycle.

GETCNT - returns the current number stored in the high speed counter.

The lowest level integrator and lineariser circuit sampling subprograms are :

GETLIS - samples a specified channel list and converts the returned character string to floating point numbers.

DVM - repeatedly samples a specified analog input channel voltage and converts the returned character string to a floating point number. Voltages are scaled by a specified Dataporte analog input attenuation factor and displayed on the terminal screen beginning at a location specified by the calling subprogram so that the display resembles a digital voltmeter (DVM). At the outset the operator is prompted for unsolicited input ('<CR>=End DVM') to terminate the process. The message '?ACCEPT?' is printed to which a response of 'Y' will cause return to the calling module. Any other response will restart the sampling process.

For every command issued the Dataporte reply is compared to the expected response to ensure that everything is functioning correctly. Any differences between actual and expected replies are examined by each module for evidence of anticipated error conditions. As well as performing a primary function many of these lowest level subprograms also perform a number of redundant circuit board fault checking procedures. For example RESET checks that the integrating status line is not activated before proceeding to activate the reset control line. Each main program uses simple graphic displays to represent the front panel layout of the instruments. All information is superimposed on these displays for the operator's convenience. Each time that a subprogram is executing an identification and an abbreviated description of it's primary function are also included on the display. Although this slows down the display cycle time and gives it a 'busy' appearance the technique is useful since if an error condition is detected (e.g. a voltage overload) or a fault (e.g. the integrating status line is not high after the integrator control line has been activated) then an abbreviated diagnostic message and possible cause and remedy are further printed on the display. Since continuous monitoring of an automated data acquisition system is boring for the most part it is envisaged that the operator may be briefly preoccupied elsewhere e.g. on another terminal and plotting previously collected results or monitoring the hot-wire signals on an oscilloscope. In order to set the operator's attention upon the detection of an error or fault the display terminal bell is made to ring periodically at a rate that depends on the seriousness of the problem. The bell continues to ring until the operator acknowledges the problem by selecting the appropriate remedial action from a displayed menu.

At an intermediate level subprograms are used to call on the lowest level subprograms to perform commonly used functions. For example INTOUT calls GETLIS with the appropriate channel list for the integrator outputs and scales the voltages with the appropriate Dataporte attenuation scale factors After storing the voltages in COMMON arrays INTOUT superimposes the numbers in the appropriate locations on the terminal graphic display. Similar functions are performed by INTINP (for circuit inputs) and INTRMS (for the RMS amplifiers and muitiplier/squarers). As an example of an intermediate level monitoring subprogram FINTST determines if the integrator finished status line is set and obtains the current high speed counter value for calculation and display of elapsed integration times.

At the highest level modules call the lower level subprograms to perform frequently used sequences of functions. The most commonly used high level subprogram used to control the integrators is INTRUN which resets the integrators, grounds the inputs to the integrators so that the offset voltages can be measured and then performs an integration cycle. A loop is entered during the integration cycle in which the current integrator outputs are sampled and updated on the graphic display, the elapsed integration time is measured and displayed, the finished status line and the unsolicited input (operator intervention) event flag are checked. The loop cycle time is around 1.5 seconds so the display is reasonably dynamic. If the operator has intervened by typing a character before the integration cycle is completed the loop is halted and a menu is presented with various options. Of course the integrators are not interrupted unless requested by the operator so that the accidental typing of a character temporarily halts the display only. After completion of the integration cycle the final outputs of the integrators are samples, stored and displayed and the terminal bell is rung once as an indicator before returning to the calling program.

6.2 Programs dedicated to performing special functions.

6.2.1 Program TUNEUP

Although the temperature sensitivity of analog circuits has reduced considerably in recent years there still remains the need to periodically tune the circuits. The offset voltage drift of the closed-loop operational amplifiers is negligible being measured in tens of microvolts per degree centigrade. After their initial adjustment the amplifier offset voltages rarely need to be adjusted again even after seasonal variations of temperature have occurred. However the multipliers are particularly prone to offset voltage drift and regularly require adjustment. The effects of relatively large offset voltages are negligible since they are measured and taken into account prior to all integrations. It is also necessary to periodically adjust the input bias current of each integrator since this is responsible for drift when the integrators are finished or placed on hold. This effect is minimised since it is automatically allowed for if a circuit calibration is performed prior to an experimental run. Nevertheless it is not good practice to allow the offset voltages to grow to large values (e.g. 10 millivolts) or to allow the input bias currents to rise to such levels as to cause integrator drift larger than around 0.2 millivolts per second.

Program TUNEUP has been specifically written to act as a step by step guide for tuning the integrator and lineariser circuits. Tuning is accomplished by adjusting circuit board mounted potentiometers which are accessible from the rear of the chassis and which are clearly labelled on the rear panels. The devices which may require tuning are divided into five groups which are displayed on the terminal screen in the form of a main tabular selection menu (see figure 7). The operator is also reminded of the correct front panel switch setting for this procedure. Once a particular group has been designated for tuning the screen is cleared and a tabular display is presented with columns corresponding to the potentiometer label, the Dataporte channel number to which the device is connected, the output voltage and the current sampling status. All device outputs are then sampled and displayed and the message 'sampled' is placed in the sampling status column. The operator then has the choice of adjusting a particular device or clearing the screen and returning to the main menu to select another group of devices or exiting from the program. If a particular device is selected for tuning its output voltage is repeatedly sampled and displayed in the appropriate tubular position so that the display has an appearance resembling that of a digital voltmeter (i.e. subroutine DVM). The message '<CR> to end DVM' is written in the sampling status column to indicate to the operator that typing a carriage return will halt the process when the output voltage has been reduced to an acceptable level. The last value sampled remains on the screen to serve as a record and the message 'adjusted' is written in the sampling status column. Figure 8 is a photograph of the screen display used for adjusting the amplifier, squarer and multiplier offset voltages. It should be noted that it is not necessary for any device to be tuned. Therefore TUNEUP can be used to check that the Dataporte is functioning correctly and that the current state of the circuits is within acceptable limits. *It is highly recommended that this be performed prior* to each session of usage.

6.2.2 Program LINSET

The cubic calibration polynomials are the result of analytical integration of the least squares curve fitting applied to the hot-wire system small perturbation sensitivities which have been determined directly from the dynamic calibration technique. In regions where the flow velocity perturbations are small compared to the mean velocity it is justifiable to assume a linear relationship between the velocity and voltage perturbations. In these regions only rms voltages need be found from which the rms velocities can be inferred. The assumption of hot-wire linearity breaks down in regions where the velocity perturbations are large. In these regions it is necessary to take the calibration nonlinearity into account. This is the function of the lineariser.

Program LINSET has been designed to act as a step by step guide for setting the two lineariser channels. Since the outputs of the lineariser channels are designed to lie between ±10 volts scaling factors may be required. These scaling factors are taken into account by LINSET. It is important to reduce the offset voltages of the lineariser to acceptable levels before proceeding since these are not measured and taken into account during the setting up procedure. The operator is required to perform the same offset tuning procedure as that prescribed in TUNEUP before proceeding with the set up procedure.

For each lineariser channel the polynomial coefficients are individually set by adjusting the appropriate front panel potentiometers which voltage divide the output of the DC voltage source/amplifier/squarer/multiplier which represent the cubic calibration polynomial terms of different order. All other inputs to the final summing amplifier must be grounded by the front panel switches while the adjustment is performed. One of two screen display files which resemble the front panel layout of each lineariser channel is displayed on the terminal and at appropriate positions on the display instructions and voltages are superimposed to provide the necessary guidance during each stage of the set up procedure. Firstly the operator is requested to connect suitable DC voltages to the lineariser input channel. The input voltage is repeatedly sampled and displayed within a labelled box on the screen until a suitable value has been adjusted whereupon a carriage return may be typed to halt the procedure. The last value measured remains on the screen as a record. Precise adjustment of this voltage is not necessary but it must remain fixed throughout the rest of the set up procedure. The next series of steps require setting of each 'coefficient' potentiometer in turn so that outputs corresponding to those calculated from the input voltage and the calibration polynomial can be adjusted. Failure to switch all the other potentiometers to ground during each adjustment procedure will produce erroneous results. The correct switch settings are flashed on and off on the screen display at each stage to emphasise their importance. While the potentiometer is being adjusted the output of the lineariser channel is repeatedly sampled and displayed within a labelled box for the operator's convenience until a carriage return is typed. Each of these adjustments should be made with precision (i.e. the calculated and adjusted value should agree with each other to within a millivolt). A tabular display is used as a record of the adjusted and calculated output voltages. Once the potentiometers have been set in sequence, any one or all of the steps may be repeated before proceeding to the next channel or exiting from the program. Figure 9 is a photograph of the screen display used for setting the V channel lineariser.

6.2.3 Integrator circuit calibration programs

Automatic integrator circuit calibration is accomplished by running one of a number of specially written programs which produce ASCII output files documenting all the circuit parameters and measured voltages that are required for the scaling and reduction of subsequent measurements. The time, date, program and operator identification codes and an operator entered comment line are also included in the output file for convenient reference purposes. Three circuit calibration programs have been written :

(a) FULCAL : which performs a full 16 integration cycle automated circuit calibration i.e. four full quadrant squarer/multiplier reference voltages are used for each gain level of the RMS integrator input amplifiers.

(b) QIKCAL : which performs a 4 integration cycle calibration i.e. one set of reference voltages per gain level.

(c) DCCAL : which performs a 1, 2, 4, 8 or 16 integration cycle calibration for the DC integrators only.

RSX11-M uses the DEC FILES11 file naming convention in which a three-letter extension and a version number are appended to the filename. Use is made of the operating system convention by automatically appending coded extensions to a single filename specified by the user i.e. this single name is the only one that need be remembered. All integrator circuit calibrations use the coded file extension 'INT'.

The autocal circuit analog switches and multiplexers are controlled by four Dataporte digital outputs whose values are set by subroutine CALMUX. Subroutine GNDINP is used to ground the inputs to the integrator circuits so that offset voltages can be recorded for subsequent data reduction. Circuit overloads and faults are monitored throughout the calibration by subroutine INTRUN as described earlier. The user is prompted to change the autocal and integrator gain levels when and if required using subroutine CGAIN. The required gain levels are flashed on and off to emphasise their importance. Figure 10 is a photograph of the terminal display. Calibration of the RMS integrators for unconventional amplifier gain settings is allowed but manual application of suitable reference voltages is required. After completion of the required number of integration cycles the results of the calibrations are displayed in tabular form which the operator is asked to examine before they are accepted and the output file is written.

6.2.4 Hot-wire calibration and run-time programs.

Programs HWCAL and XWCAL have been written to perform a number of normal and crossed-wire small perturbation sensitivity calibrations from which the full nonlinear cubic calibration polynomials are derived. Programs HWRUN and XWRUN are used to reduce real-time data. It should be remembered that the RMS integrator input amplifiers must be ac coupled during the hot-wire calibration and run-times.

All programs use a common set of subroutines. Initially each program must call subroutine CALFIL which reads the values of the circuit calibration file parameters as variables which are stored in common block. The most frequently used subprogram is subroutine RUNRED which first calls INTRUN so that the 'run-time' circuit offset voltages are measured and an integration cycle is performed. RUNRED then processes the run-time and the previously determined circuit calibration offset voltages to produce the temporal mean and rms values of the signals applied to the integrator circuits. When the small perturbation sensitivity calibration operating point is changed or the run-time position of the probe is altered the need may arise to alter the gain levels of the RMS integrator input amplifiers. Subroutine RGAIN is a small subprogram designed to superimpose on the integrator from panel graphic display the gain levels that are currently being assumed for the data reduction. The gain levels are flashed on and off to emphasise their importance. Of course it is essential that the gain levels indicated on the display and those selected on the front panel correspond. A typed response of any other character than 'Y' in response to the prompt 'Change gains?' printed by RGAIN will return control to the calling program. Otherwise new gain levels can be entered in response to a series of prompts.

During HWCAL and XWCAL the small perturbation sensitivity is evaluated at operating points which are varied by altering the tunnel free-stream velocity which is measured with a Pitot-static tube attached to a differential manometer. Subroutine PITVEL has been written for the operator's convenience during this adjustment procedure. The manometer output is repeatedly sampled, converted to the tunnel velocity and displayed at the same location on the terminal display as the tunnel motor speed is adjusted. When a satisfactory speed has been obtained the operator types a carriage return to end the process. The mean velocity used for the nonlinear calibration polynomial is determined from the integrated value of the manometer output. The 120 pulse/cycle signal from the optical encoder on the calibrator is used as the input to a low speed counter so that the (fractional) number of calibrator cycles that occurred during the integration time period can be obtained. RUNRED first calls INTRUN which returns the unscaled integrator voltages and the integration time period so that the rms velocity perturbation can be calculated. RUNRED then scales the voltages so that the small perturbation sensitivities can be calculated in the main program.

At the end of the required number of small perturbation sensitivity evaluations the results are displayed on the terminal in tabular form. The full nonlinear calibration polynomials are listed together with the lineariser 'coefficient' potentiometer settings for +10 volts reference inputs. The tables list the measured voltages, velocities and sensitivities and those calculated from the curve fits together with absolute and percentage deviations. Tables 1(a) and (b) are hardcopy printouts of calibration data from XWCAL for a typical crossed-wire probe. In table 1(a) the matched outputs E_u and E_v have been applied directly to the integrators for calibration while in table 1(b) the results obtained by using the lineariser are shown. As would be expected the small perturbation sensitivities of the lineariser outputs show only a slight variation with free-stream velocity. In both cases the second-order polynomial fits to the sensitivities agree closely with the measured data. The products of the sensitivities $\partial U/\partial E_u$ and $\partial V/\partial E_v$ provide a cross-check for the Reynolds stress sensitivities $\partial(UV)/\partial(E_u E_v)$ which have been measured directly and it is apparent that the agreement is excellent. The unknown constant of integration A for the calibration polynomials has been obtained by bodily shifting the curve obtained from the sensitivities to obtain the least square deviation from the mean values. The sensitivities have been evaluated on the bases of the rms velocity which is calculated from the measured calibrator period and stroke. The mean velocities have been measured with a Pitot-static tube. Therefore the excellent agreement of the calibration polynomial with the mean data points gives added confidence in the method since two independent techniques have been used to measure velocity. The operator is asked to examine the results before they are accepted. Accepted hot-wire calibrations are written in an ASCII file containing all the circuit parameters, voltages, mean calibrator periods, gain levels, integration periods needed for each small perturbation sensitivity evaluation. The date and time of calibration, operator and program identification codes and an operator entered comment line are also included in the output file. The raw data are written so that the full nonlinear calibration polynomials can be recalculated by HWRUN and XWRUN. The raw data are written (rather than just the calibration polynomial coefficients) so that a log of the process is recorded. The coded file extensions for the calibration files are 'HWC' and 'XWC' for HWCAL and XWCAL respectively.

The programs HWRUN and XWRUN read the circuit calibration data file (using CALFIL) and the hot-wire calibration file data (using subroutines HWCFIL or XWCFIL) as variables which are stored in COMMON blocks. Run data are measured by RUNRED to give the mean and rms voltages which are converted into mean and rms velocities by subroutines HWRED and XWRED. There is no need for the programs to differentiate between matching circuit or lineariser outputs since the same techniques are used for both. After each measurement the operator has the choice of clearing the screen and examining the results collected so far. The probe is then traversed to a new position and RGAIN is called to enquire whether the current gain settings need to be changed. The cycle is repeated until the predetermined number of measurements have been performed. After a table of processed results has been displayed and accepted, the output files are written. Again all details of the run-time operations are recorded in the output files (rather than just the final results) so that a log is produced. The coded file extensions for run data are 'HWD' and 'XWD'.

7. CONCLUDING REMARKS

RSX11M provides a service to users called the Resource Monitor Display (RMD) in which the current status of the system is displayed and updated at one second intervals. Included in the display is the task that is currently executing in the CPU. Over a long period of time the display in effect provides a stimate of the proportion of CPU real-time that a task is using. The RMD display was observed over a 15 minute period while the hybrid data acquisition system was functioning with heavy RS232C activity and not once did the task appear as executing. This is because all input/output activity occurs independently of the CPU. The transparency of the hybrid data acquisition software to the operating system means that many other tasks can run concurrently without delays. The hot-wire system is still undergoing development. So far the programs have been small enough to not require the multi-tasking capabilities.

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File=XWMATCH, User=JHW, Started 09-DEC-85, 09:34:54
Simultaneous test run, calibration of output of matching circuit
Temp=21.6 deg. C, Density = 1.19534, Viscosity = 0.1537E-04

Crossed-wire Calibration

dU/dEu	=-0.5924E-03(Eu**2)	0.3343E-01 (Eu)	0.5068E+00
dV/dEv	=-0.6835E-03(Eu**2)	0.3058E-01 (Eu)	0.5114E+00
d(u*v)/d(eu*ev)	=-0.1636E-02(Eu**2)	0.3329E-01 (Eu)	0.2608E+00

U= 5.630		0.5086E+00(Eu)	0.1672E-01(Eu**2)	0.1975E-03(Eu**3)
V= 0.067	0.7831E-02(Eu)	0.5114E+00(Ev)	0.3058E-01(Eu*Ev)	0.6835E-03(Eu**2*Ev)

N	Period sec	U m/s	u'/U %	Eu volts	Ucalc m/s	Vcalc m/s	ALPHA degrees	U-Ucalc m/s	Ucalc/U %
1	0.429	7.868	4.989	3.810	7.821	-0.002	-0.013	-0.046	-0.589
2	0.429	7.137	5.500	2.654	7.101	-0.003	-0.022	-0.036	-0.501
3	0.429	6.509	6.030	1.623	6.500	-0.002	0.018	-0.009	-0.141
4	0.430	6.064	6.458	0.823	6.060	0.005	0.044	-0.004	-0.068
5	0.429	5.363	7.318	-0.518	5.317	0.004	0.043	0.007	0.135
6	0.426	4.768	8.290	-1.742	4.793	0.002	0.022	0.025	0.533
7	0.427	4.331	9.105	-2.179	4.366	0.000	-0.004	0.035	0.818
8	0.425	3.960	10.006	-3.620	3.998	0.000	-0.004	0.038	0.967
9	0.424	3.516	11.296	-4.774	3.561	-0.004	-0.072	0.045	1.287
10	0.424	7.843	5.063	3.757	7.787	-0.003	-0.022	-0.056	-0.716

	dU/dEu (m/s)/v	dV/dEv (m/s)/v	d(u*v) d(eu*ev)	dU/dEu (m/s)/v	dV/dEv (m/s)/v	d(u*v) d(eu*ev)	dU*dV dEu dEv	(Su)m (Su)c	(Sv)m (Sv)c	(Suv)m (Suv)c
	—MEASURED VALUES—			—CURVE FIT VALUES—			CHECK	%	%	%
1	0.644	0.639	0.412	0.645	0.638	0.411	0.412	-0.04	0.13	0.27
2	0.602	0.599	0.361	0.602	0.597	0.361	0.360	0.02	0.18	0.02
3	0.565	0.563	0.319	0.564	0.567	0.319	0.318	0.06	0.07	-0.18
4	0.537	0.538	0.289	0.537	0.537	0.289	0.289	0.11	0.15	0.00
5	0.493	0.496	0.245	0.491	0.496	0.244	0.245	0.23	0.13	0.34
6	0.450	0.458	0.207	0.452	0.460	0.208	0.206	-0.48	-0.38	-0.53
7	0.422	0.433	0.183	0.422	0.433	0.182	0.183	0.07	-0.09	0.43
8	0.395	0.409	0.162	0.395	0.410	0.162	0.162	-0.15	-0.13	0.03
9	0.363	0.382	0.139	0.363	0.381	0.139	0.139	0.22	0.28	0.15
10	0.642	0.634	0.408	0.643	0.636	0.409	0.407	-0.05	-0.34	-0.24

TABLE 1a CALIBRATION DATA FROM XWCAL -
TYPICAL CROSSED-WIRE PROBE -
MATCHED OUTPUT APPLIED DIRECTLY TO INTEGRATORS

File=XWLIN, User=JHW, Started 09-DEC-85, 09:34:54
Simultaneous test run, calibration of output of lineariser
Temp=21.6 deg. C, Density = 1.19534, Viscosity = 0.1537E-04

Crossed-wire Calibration

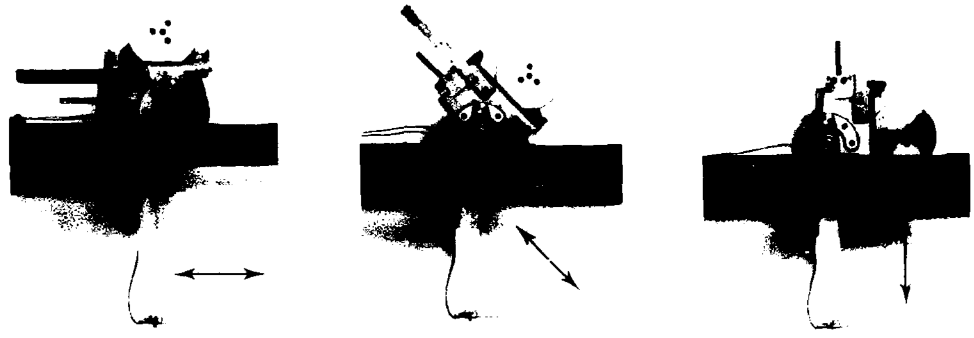
dU/dEu	=-0.1519E-03(Eu**2)	0.1840E-01 (Eu)	0.9502E+00
dV/dEv	=-0.1071E-03(Eu**2)	0.2112E-01 (Eu)	0.9136E+00
d(u*v)/d(eu*ev)	=-0.2588E-02(Eu**2)	0.3979E-01 (Eu)	0.8634E+00

U= 0.066	0.9502E+00(Eu)		0.9199E-02(Eu**2)	-0.5064E-03(Eu**3)
V=-0.010	-0.1984E-02(Eu)	0.9136E+00(Ev)	0.2112E-01(Eu*Ev)	-0.1071E-02(Eu**2*Ev)

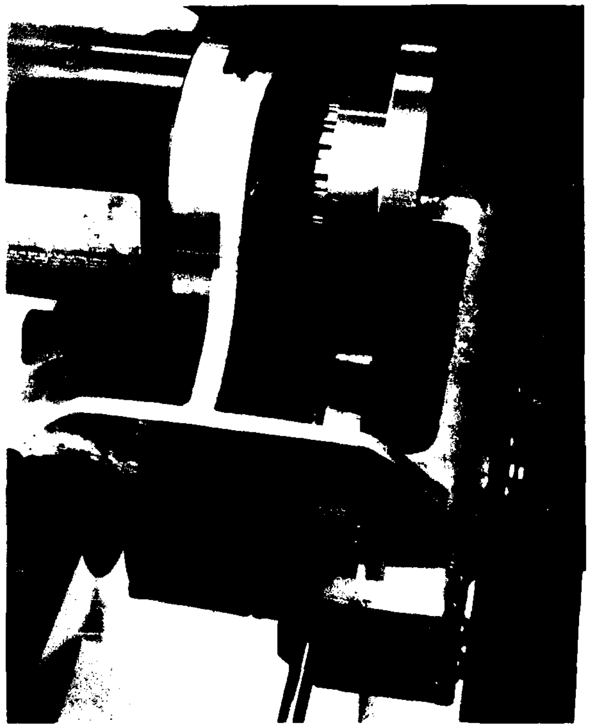
N	Period sec	U m/s	u'/U %	Eu volts	Ucalc m/s	Vcalc m/s	ALPHA degrees	U-Ucalc m/s	Ucalc/U %
1	0.433	7.868	4.943	7.822	7.819	-0.003	-0.025	-0.048	-0.616
2	0.431	7.137	5.474	7.114	7.109	0.001	0.011	-0.028	-0.387
3	0.430	6.509	6.016	6.502	6.494	0.000	-0.001	-0.015	-0.228
4	0.430	6.064	6.458	6.068	6.057	0.004	0.040	-0.007	-0.108
5	0.429	5.363	7.318	5.369	5.369	0.006	0.060	0.006	0.104
6	0.431	4.768	8.194	4.803	4.786	0.002	0.030	0.018	0.372
7	0.427	4.331	9.105	4.385	4.367	-0.001	-0.011	0.036	0.833
8	0.426	3.960	9.982	4.011	3.993	-0.003	-0.041	0.033	0.837
9	0.424	3.516	11.296	3.581	3.564	-0.004	-0.059	0.048	1.377
10	0.423	7.843	5.075	7.803	7.800	-0.003	-0.020	-0.043	-0.554

	dU/dEu (m/s)/v	dV/dEv (m/s)/v	d(u*v) d(eu*ev)	dU/dEu (m/s)/v	dV/dEv (m/s)/v	d(u*v) d(eu*ev)	dU*dV dEu dEv	(Su)m (Su)c	(Sv)m (Sv)c	(Suv)m (Suv)c
	—MEASURED VALUES—			—CURVE FIT VALUES—			CHECK	%	%	%
1	1.003	1.015	1.021	1.001	1.013	1.016	1.019	0.23	0.20	0.44
2	1.003	1.008	1.012	1.004	1.010	1.015	1.011	-0.16	-0.16	-0.31
3	1.004	1.005	1.011	1.006	1.006	1.013	1.009	-0.17	-0.03	-0.19
4	1.007	1.005	1.013	1.006	1.002	1.010	1.012	0.15	0.23	0.38
5	1.004	0.995	0.999	1.005	0.996	1.003	0.998	-0.16	-0.17	-0.33
6	1.007	0.992	0.999	1.003	0.990	0.995	0.998	0.30	0.14	0.45
7	1.001	0.984	0.986	1.002	0.986	0.988	0.986	-0.04	-0.13	-0.17
8	1.000	0.981	0.982	1.000	0.981	0.981	0.982	0.07	0.04	0.09
9	0.995	0.976	0.972	0.997	0.975	0.973	0.971	-0.13	0.00	-0.12
10	1.000	1.012	1.014	1.001	1.013	1.016	1.012	-0.09	-0.14	-0.25

TABLE 1b CALIBRATION DATA FROM XWCAL - TYPICAL CROSSED WIRE PROBE - OUTPUT FROM LINEARISER

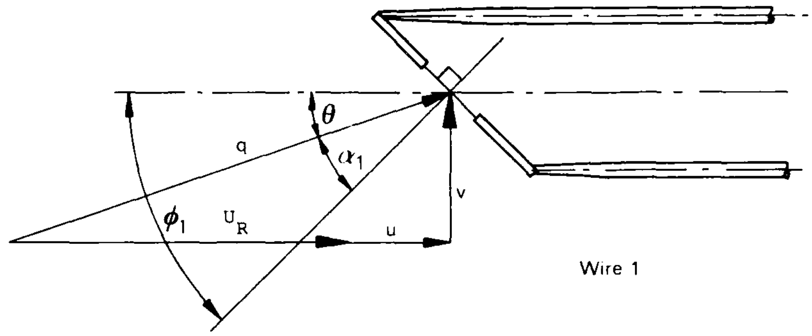


(a) (b) (c)



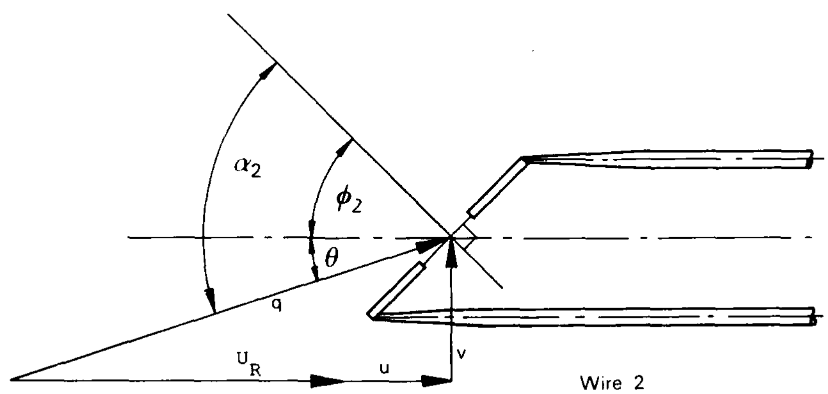
(d)

FIGURE 1. THE DYNAMIC CALIBRATOR WITH STING AND CROSSED-WIRE PROBE MOUNTED. TOP PHOTOGRAPHS (a), (b) AND (c) ARE SIDE ON VIEWS SHOWING THE CALIBRATOR SET UP FOR HORIZONTAL, FORTY-FIVE DEGREE AND VERTICAL SHAKING RESPECTIVELY. OBLIQUE VIEW (d) SHOWING MECHANICAL DETAILS.



$$\cos \alpha_1 = \cos \phi_1 \cos \Theta + \sin \phi_1 \sin \Theta$$

$$E_1{}^2 = A_1 + B_1 \sqrt{q \cos \alpha_1} = A_1 + B_1 \sqrt{(U_R + u)\cos \phi_1 + v \sin \phi_1}$$



$$\cos \alpha_2 = \cos \phi_2 \cos \Theta - \sin \phi_2 \sin \Theta$$

$$E_2{}^2 = A_2 + B_2 \sqrt{q \cos \alpha_2} = A_2 + B_2 \sqrt{(U_R + u)\cos \phi_2 - v \sin \phi_2}$$

Figure 2. NOTATION FOR CROSSED HOT WIRES.

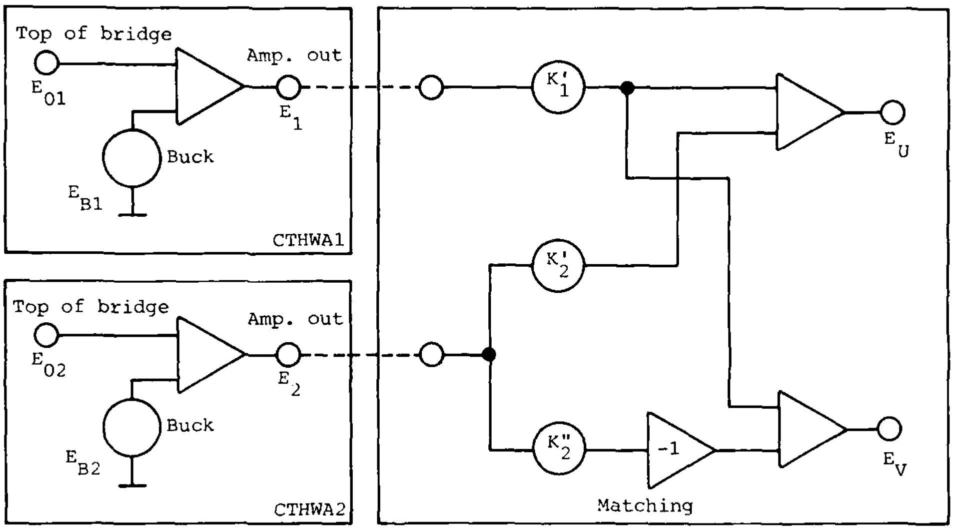


FIGURE 3. CROSSED-WIRE MATCHING CIRCUIT.

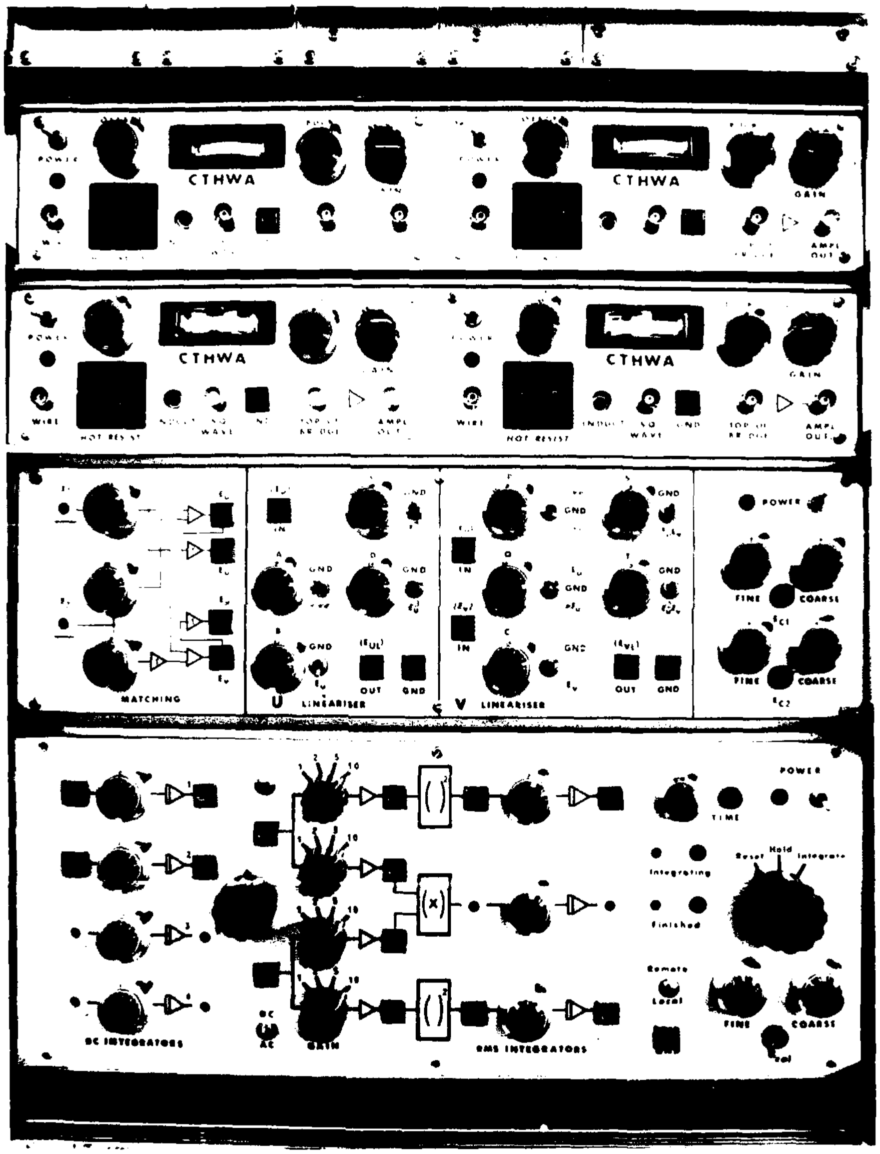


FIG. 4. CLOSE UP PHOTOGRAPH OF PORTABLE RACK SHOWING (FROM TOP TO BOTTOM) TWO DUAL CHANNEL HOT-WIRE ANEMOMETERS, THE LINEARISER AND MATCHING CIRCUITS AND THE INTEGRATORS.

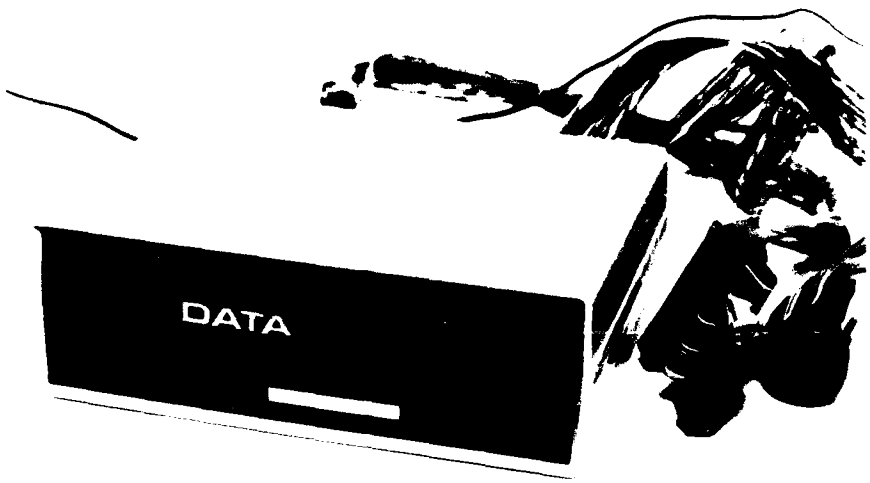


FIG. 5. PHOTOGRAPH OF MICROPROCESSOR BASED DATA ACQUISITION SYSTEM (DATAPORTE) WITH 46 ANALOG INPUT AND 8 DIGITAL INPUT/OUTPUT CHANNELS USED FOR SAMPLING, MONITORING AND CONTROLLING THE INTEGRATORS AND LINEARISER CIRCUITS. THR DATAPORTE IS REMOTELY CONTROLLED BY A PDP 11/23-PLUS MICROCOMPUTER VIA AN RS232C SERIAL INTERFACE.

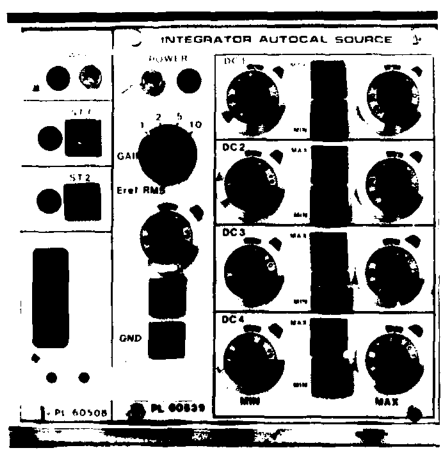


FIG. 6. CLOSE UP PHOTOGRAPH OF INTEGRATOR AUTOCAL SOURCE FOR AUTOMATED INTEGRATOR CIRCUIT CALIBRATION.

CIRCUIT TUNEUP

NOTE: 1. REQUIRES ACCESS TO REAR OF INTEGRATORS AND LINEARISER.
2. ALL FRONT PANEL INPUTS MUST BE PATCHED TO GROUND.

	M E N U	NO. TIMES
1	LINEARISER OFFSETS	1
2	AMP.,SQ. & MULT. OFFSETS	1
3	DC INTEGRATOR OFFSETS	1
4	RMS INTEGRATOR OFFSETS	1
5	INTEGRATORS' BIAS CURRENT	0

<CR>=Exit, or Select from <1,2,3,4,5>, Choice=

FIG. 7. TYPICAL VT100 TERMINAL BOX GRAPHICS DISPLAY. SELECTION MENU SHOWN FOR GROUPS TO BE SELECTED FOR OFFSET VOLTAGE ADJUSTMENT.

AMPLIFIER, SQUARER & MULTIPLIER OFFSETS

	POT LABEL	TEST POINT	OFFSET (MV)	CURRENT STATUS	DP. CHANNEL
STEP 1.	ACDC1	TST (R)	0.105	ADJUSTED	9
STEP 2.	ACDC2	TST (R)	0.078	ADJUSTED	9+
STEP 3.	GSQ1	FRONT P.	0.045	ADJUSTED	19-
STEP 4.	GMU1	FRONT P.	0.113	ADJUSTED	20
STEP 5.	GMU2	FRONT P.	0.098	ADJUSTED	20+
STEP 6.	GSQ2	FRONT P.	0.052	ADJUSTED	21-
STEP 7.	SQ 1	OFF (R)	0.127	CR =END DVM	21+
STEP 8.	MULT	OFF (R)	1.347	SAMPLED	22-
STEP 9.	SQ 2	OFF (R)	2.061	SAMPLED	22+

CR =End or Steps 1...9 or 10 =All steps, Choice=

FIG. 8. DISPLAY FOR RMS INTEGRATOR AMPLIFIERS, SQUARER AND MULTIPLIER OFFSET ADJUSTMENTS. ALL OFFSETS HAVE BEEN SAMPLED AND THE OPERATOR HAS ADJUSTED THE FIRST 6 CHANNELS OF THE GROUP. THE OUTPUT OF THE SQUARER IS BEING REPETITIVELY SAMPLED AND DISPLAYED AT THE SAME LOCATION ON THE SCREEN SO THAT THE NUMBER TAKES ON THE SAME APPEARANCE OF A DIGITAL VOLTMETER.

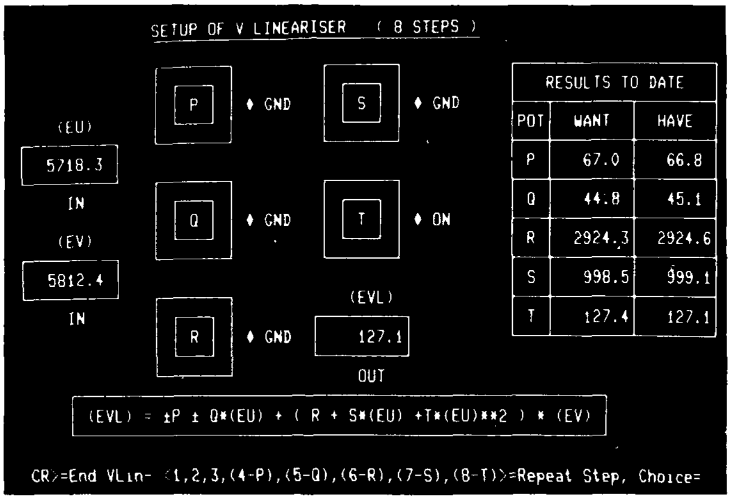


FIG. 9. TERMINAL DISPLAY RESEMBLING THE FRONT PANEL LAYOUT FOR THE SETTING OF THE V CHANNEL LINEARISER (SEE FIG. 4.). THE OPERATOR HAS JUST FINISHED THE EIGHT STEPS AND HAS THE CHOICE OF REPEATING ANY ONE. THE LABELS 'GND' AND 'ON' FLASH ON AND OFF TO INDICATE THE CORRECT FRONT PANEL SWITCH SETTINGS. (VOLTAGE IN MILLIVOLTS).

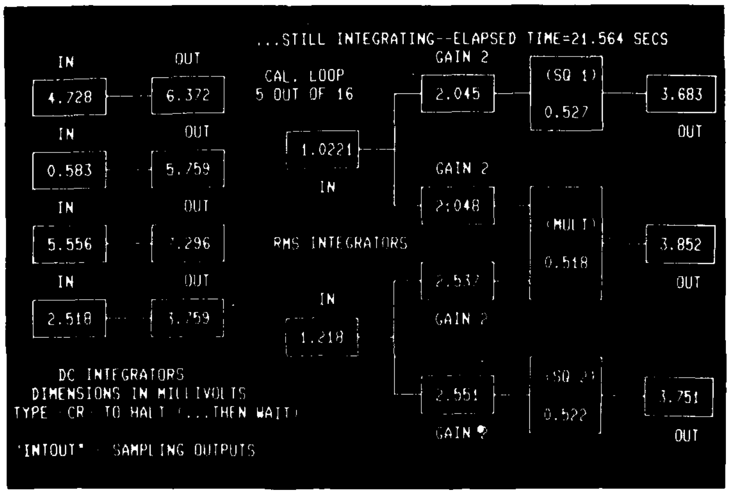


FIG. 10. TERMINAL DISPLAY RESEMBLING THE FRONT PANEL LAYOUT OF THE INTEGRATORS (SEE FIG. 4.). THE NUMBERS IN THE LABELLED BOXES CORRESPOND TO THE VOLTAGES MEASURED AT THE SOCKET LOCATIONS ON THE FRONT PANEL. ELAPSED TIME AND CURRENT PROGRAM FUNCTIONS ALSO DISPLAYED.

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AL 149
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PAGE CLASSIFICATION
UNCLASSIFIED

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1a. AR NUMBER	1b. ESTABLISHMENT NUMBER	2. DOCUMENT DATE	3. TASK NUMBER
AR-004-509	ARL-AERO-R-172	NOVEMBER 1986	DST 82/021

4. TITLE
A HYBRID HOT-WIRE DATA ACQUISITION SYSTEM

5. SECURITY CLASSIFICATION
(PLACE APPROPRIATE CLASSIFICATION IN BOX(S) IE. SECRET (S), CONFIDENTIAL (C), RESTRICTED (R), UNCLASSIFIED (U).)

DOCUMENT	TITLE	ABSTRACT
U	U	U

6. No. PAGES
45

7. No. REFS.
10

8. AUTHOR(S)
J.H. WATMUFF

9. DOWNGRADING/DELIMITING INSTRUCTIONS

10. CORPORATE AUTHOR AND ADDRESS

AERONAUTICAL RESEARCH LABORATORIES
P.O. BOX 4331, MELBOURNE VIC. 3001

11. OFFICE/POSITION RESPONSIBLE FOR

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14. DESCRIPTORS
Hot wire anemometers
Data acquisition
Hybrid systems

15. DRDA SUBJECT CATEGORIES
14020
01010

16. ABSTRACT
This report describes the philosophy, software, hardware and operation of a hybrid hot-wire signal processing system whcre unscaled mean and rms quantities are determined by analog circuits whose output voltages are passed to a digital computer for scaling and further processing. The circuits are orchestrated by a compact intelligent microprocessor based data acquisition system which can be supervised by any computer with an RS232C serial interface.

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16. ABSTRACT (CONT.)

17. IMPRINT

AERONAUTICAL RESEARCH LABORATORIES, MELBOURNE

18. DOCUMENT SERIES AND NUMBER	19. COST CODE	20. TYPE OF REPORT AND PERIOD COVERED
Aerodynamics Report 172	556055	

21. COMPUTER PROGRAMS USED

22. ESTABLISHMENT FILE REF. (S)

23. ADDITIONAL INFORMATION (AS REQUIRED)