Portable monitoring instrumentation
for use on animals under anesthesia

by

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INTRODUCTION

The need for intensive care monitoring has been recognized for centuries, as the continuous monitoring of a patient's status enables problems to be detected as they occur, and proper emergency aid administered before permanent damage results. In the monitoring of small animals - i.e., dogs and cats - in the clinical setting, it is often unnecessary to have the accuracy and hard-copy recording of a chart recorder, with its associated inconveniences of calibration and calculations. It is often more desirable to have modular monitoring devices with digital readouts and limit detectors, which avoid the necessity of having to continually monitor the recordings and perform the appropriate calculations.

It is therefore the objective of this research project to design devices that monitor the following physiological parameters:

- Heart rate
- Respiration rate
- Core temperature
- Arterial blood pressure

and have the following design criterion:

- Portability - battery operation
- Ease of use
- Digital readout

- Variable limit detection
- Audible and visual warning signals
- Battery level tester
- Low cost
- Sturdy packaging

All monitoring devices are designed using ready-made low-powered integrated circuit subsections, and as few 9 V batteries as possible.

LITERATURE REVIEW

Physicians do not always agree among themselves as to which physiological parameters should be monitored, with the number monitored weighed against cost, complexity, and reliability. Intensive monitoring and intensive treatment presents a complex problem of information analysis, with the patient's condition followed carefully by repeated measurements of many variables, and treatment prescribed based on the results obtained. Critical patient areas requiring a high level of sophisticated equipment include the operating room, where surgical procedures are performed; the intensive care unit, where post-surgical followup is conducted and where medical patients with very serious problems are treated; and the coronary care unit, where heart attack victims are observed (Cromwell et al., 1980). In all of these areas, heart rate, arterial blood pressure, respiration rate, and body temperature constitute a common segment of the array of instruments making up a patient's bedside station, with all equipment being of the dynamic monitoring type, delivering continuous quantitative indications (Jacobson and Webster, 1977).

The Monitoring of Heart Rate

Since the body can go into circulatory failure in a short time, automatic monitoring techniques are important. By measuring the arterial blood pressure, central venous pressure, and ECG, imminent circulatory failure can be detected and treatment quickly begun (Jacobson and Webster, 1977). The heart rate for humans is in the range of 45 to 200 beats per minute, while that for laboratory animals is in the range of 45 to 600 beats per minute (Furno, 1981a; Thomas, 1974).

A cardiotachometer is an integral part of a patient's monitoring equipment, dynamically monitoring the patient's heart or pulse rate and concurrently providing a warning when a significant change occurs (Hinson and Wilkinson, 1977). Its function, electronically, is to convert a preconditioned physiological signal into a voltage proportional to the patient's heart rate. This voltage is then displayed on a calibrated panel meter, often digital, to increase the accuracy and ease of reading. Alarm circuitry is usually provided, designed to trigger an audible and visual alarm when the patient's heart rate drops below or exceeds preset limits, being especially useful when a nurse is unable to keep constant watch on the cardioscope (Jacobson and Webster, 1977). Thus, the front panel of a cardiotachometer usually includes the rate reading on a meter, adjustable high/low alarms, alarm lights, a lamp indicating detection of a beat or pulse, and a sensitivity control (Thomas, 1974).

Methods of transduction

Although pulse rate can be easily determined manually by counting the number of pressure pulses of the radial artery at the wrist, this physiological characteristic is not normally used to trigger an electronic rate meter (Furno, 1981b). The signals most commonly used include:

- Some identifying feature of the ECG, most commonly, the QRS complex,
- The arterial pressure waveform, which is particularly useful in the presence of electrocautery interference or pacemaker artifacts, as they do not cause false triggering of the ratemeter (Fox, 1977), and
- The pulses produced by a photocell attached to the finger or ear lobe, which responds to the very small decrease in light transmission each time the heart pumps blood through the capillaries (Plant, 1978).

The mechanical work of the heart is accompanied by electrical activity which always slightly precedes muscle contraction (Cromwell et al., 1980; Furno, 1981b; Jacobson and Webster, 1977; Spooner, 1977a). The waveform of the ECG is thus produced by the rhythmic changes in the direction and magnitude of the vector sum of the dipoles created by a depolarization wave travelling along the individual muscle cells associated with the contraction of the atria and the ventricles. To measure these bioelectric potentials, a transducer capable of converting ionic potentials and currents into electric potentials and currents is required (Cromwell et al., 1980; Furno, 1981b). Such a transducer consists of two electrodes which measure the ionic potential difference between their respective points of application. In ECG recordings, the low conductivity of the outermost horny layer of the skin presents the greatest problem; thus, by making direct contact with the subdermal tissue or the intercellular fluids, needle electrodes and other types of electrodes that create an interface beneath the surface of the skin, have lower electrical impedances and are also less susceptible to motion artifact (Cromwell et al., 1980; Jacobson and Webster, 1977).

Amplification of the ECG signal from its low values on the body's surface, peak range 10 uV to 5 mV (Thomas, 1974), to levels large enough to be used by the instrumentation circuitry, is performed by a special purpose differential amplifier. A block diagram of a typical ECG amplification system would include:

- Protection circuit - Blocks the high voltages induced by electrosurgical and defibrillation units, with slew rate limiters reducing the pacing or defibrillator pulses when present in the ECG input (Furno, 1981b; Thomas, 1974).

- Isolation - Prevents currents greater than 10 uA from flowing through the patient (Furno, 1981b).
- Differential amplifier - Amplifies the ECG signal while rejecting electrical interference (Furno, 1981b; Jacobson and Webster, 1977; Thomas, 1974). Typical characteristics include:

  High input impedance - greater than 10 megohms
  Good common mode rejection - -60 dB (1000:1)
  Limited bandwidth - 0.05 to 100 Hz (low-order
  time-constant = 3 sec)
- Electrode fall-off detector - Consists of a high-frequency, of the order of 50 kHz, current source passing several mA through the electrodes, detecting poor electrode connections by the changes in the electrical impedance (Furno, 1981b).

To determine the heart rate, a count of some identifying feature of the ECG must be made, usually the QRS complex. The primary task of an R-wave trigger is to discriminate the R-wave from other ECG components, accommodating a wide range of QRS complex amplitudes, 0.07 to 1.13 mV for Lead I, and durations, 50 to 100 ms (Thomas, 1974), while minimizing false triggering due to noise artifact. This problem has been approached using a variety of techniques both digital and analog, with the characteristics most often used to differentiate the R-wave from other components including amplitude, slope, frequency, sequency, and wave morphology (Taylor, 1979). For improved reliability, techniques such as squaring and linear filtering to enhance the QRS complex and blocking circuits to prevent triggering by the T-wave have been used (Taylor, 1979). Spectral analysis of the ECG signal reveals that most of the frequency contained in the QRS complex lies near 16 Hz; therefore, by using a bandpass filter, the QRS complex can be retained, while the P- and T-waves, high-frequency interference, low-frequency artifact, and baseline drift are all rejected (Furno, 1981b; Hinson and Wilkinson, 1977; Jacobson and Webster, 1977; Thakor, 1981).

The determination of rate

Frequency measurements are commonly based on the number of events occurring during a fixed interval of time. Cardiotachometers produce a voltage proportional to the count in beats per minute by either averaging a number of beats or by measuring the time interval between beats and displaying the reciprocal.

The frequency-to-voltage converter of the averaging cardiotachometer consists of a monostable multivibrator and an integrator/low-pass filter circuit (Furno, 1981b; Hinson and Wilkinson, 1977). The multivibrator serves to convert each triggering pulse received from the QRS complex detector to a pulse of constant amplitude and constant duration, one pulse per complex. The integrator/low-pass filter circuit integrates the string of positive pulses from the monostable, converting them to a DC output equal to the time averaged amplitude. Because of the tradeoff between the amount of output ripple acceptable and the required time-constant of the filter, often 5 to 15 seconds, this method has the disadvantage of being slow when very low frequencies are measured (Roy and Wehnert, 1974; van den Steen, 1979).

The measurement of the time interval between two successive events may thus be a more suitable approach to physiological rate monitoring. To convert time to frequency, the beat-to-beat cardiotachometer must compute the reciprocal value of the time interval measured, producing instantaneous, or within one beat, results (Roy and Wehnert, 1974; van den Steen, 1979). Since the changes in the time interval will appear as changes in the instantaneous value displayed, the response to changing heart rates will be much more rapid than that of an averaging cardiotachometer. However, this is accomplished at the cost of increased complexity, and for simple trend monitoring, the averaging technique is often adequate

(Furno, 1981b).

Simulating the function $f = 1/T$ is probably the most difficult part of building a low-cost rate meter. Although analog dividers are basic building blocks in a wide variety of applications, until recently they remained bulky, very limited in operating range and prohibitively expensive. Recently, however, they have profited from the kinds of technological and design advances that have characterized the progress of integrated circuits in other areas. Now, dedicated analog dividers and multifunction converters are available in dual in-line packages, and their low price has gone hand-in-hand with performance that has improved by orders of magnitude. Wong (1979) reviews the subjects of multiplier-inverted-dividers, analog dividers, and multifunction converters, comparing their characteristics and capabilities.

Parviainen et al. (1978) describe a rate meter which can determine the heart rate from both the ECG and the arterial pressure curve. The calculation of heart rate is based on the detection of a selected level in the pressure waveform or ECG, with hysteresis provided to eliminate false triggering caused by aortic notches or P-waves. The device also contains a "block" circuit to inhibit false triggering for a period proportional to the previous inter-beat interval. To determine rate, a timing pulse activates a sample-and-hold circuit, sampling the output of the integrator, which is then reset to a fixed reference voltage and allowed to start the generation of a negative-going ramp. The sampled voltage is subtracted from the integrator reset voltage to give a difference which is proportional to the period between timing pulses, with the heart rate then obtained by dividing a fixed voltage by the inter-beat interval, using an 8013 (Intersil) analog multifunction circuit in the divider configuration. The circuit described is reported to have a 2% accuracy in the range of 70 to 300 beats per minute and a 10% accuracy in the range of 40 to 500 beats per minute.

The classical method of measuring a time interval, described by Roy and Wehnert (1974), is to use a "hold" circuit with two RC networks which are alternately charged and discharged. The circuit uses the exponential relationship between capacitor voltage and charging time

$$Vc = Vb\ (1 - \exp(-t/RC))$$

with one RC network being charged, while the other, which has already been charged, holds its value for the indicator display. If a linear output is desired with respect to the pulse period, then the charge on the capacitor should be kept small in comparison with the total charge available, i.e., the time-constant should be large with respect to the period. However, if a linear relationship is desired between frequency and the output voltage, as is often the case, then the inverse of time must be linear, since f = 1/t. The approach used is to operate on a specific portion of the capacitor charging cycle where the exponential charging curve can be best approximated by a hyperbola, between 0.63 and 0.99 of the final value, i.e., between 1 and 5 charging time-constants.

Others have also taken advantage of the approximately hyperbolic relationship between voltage and time over a small portion of the capacitor charge/discharge curve, utilizing only one RC network and a sample-and-hold circuit. Each time a QRS complex is detected, a one-shot is triggered which causes the value of the RC circuit to be sampled and stored. The trailing edge of the sampling pulse then triggers another one-shot, resetting the RC network which then begins charging or discharging to allow the measurement of the next inter-beat interval. The circuit described by Purves (1975) sacrifices some linearity for reduced cost, being linear within 6% of full scale for the range 40 to 200 beats per minute. If the range is to be extended downward, with better resolution at low frequencies, the time-constant of the circuit described can be increased, but at the loss of linearity. In the circuit described by Cousin and Smith (1978), the error over the operating range of 55 to 220 beats per minute is 2 beats per minute. By the addition of a small number of components, whose operation can be described as a second exponential time-constant that begins after a preselected period has elapsed, the basic circuit described can be enhanced to provide an extended range at low rates, with the error over this extended range held to within 4 beats per minute.

van den Steen (1979) describes a method of determining rate by utilizing a monolithic multiplier circuit, though not as a multiplier-inverted-divider, as is often the case. After a "dead-time" interval, To, following the first event, the 1/t generating circuit is started with an initial voltage Vo = 1/To, corresponding to fmax = 1/To. On arrival of a second event, the output voltage of the 1/t circuit is held constant and stored by a sample-and-hold circuit. This sampling requires a portion of the dead-time mentioned above, with the remainder being used to reset the 1/t circuit to its initial value. A method for generating a voltage corresponding to the inverse function arises from consideration of the differential equation

$$\{f(t)\}^2 + c\{df(t)/dt\} = 0$$

the solution of which is the required function f(t) = c/t. This equation can be rewritten in the integral form and simulated with a monolithic multiplier circuit and analog computer elements.

The Monitoring of Respiration Rate

The purpose of respiration is to supply the body's cells with oxygen from the environment and to eliminate the carbon dioxide produced during metabolic processes (Cromwell et al., 1980; Jacobson and Webster, 1977). Although no single laboratory test or even a simple group of tests is capable of completely measuring pulmonary function, both mechanical and physiological (Cromwell et al., 1980), respiratory function can be assessed by estimating ventilation and by blood analysis (Jacobson and Webster, 1977). In practice, however, the possibilities of quantitatively monitoring ventilatory volume are limited, as the patient must breath through an apparatus which can often be disturbing, yielding data which may not be worthwhile. Thus, only respiration rate is usually monitored, with the respiration rate of humans in the range of 12 to 40 breaths per minute, and that for laboratory animals lying in the range of 6 to 60 breaths per minute (Furno, 1981a; Thomas, 1974).

Respiratory arrest is second only to cardiac arrest in its ability to cause rapid death of a hospital patient (Cooley, 1977). The unpredictable occurrence of apnea is also a problem commonly encountered during experiments with anesthetized laboratory animals (Chess et al., 1976). This condition leads to the rapid deterioration of the physiological state of the patient, and if not corrected within a few minutes, to death or neurological damage. The greatest need for spontaneous respiration monitoring exists in patients where the probability of apnea is high, such as in premature infants, thoracic surgery patients, persons with pulmonary or cardiac disease, neurosurgical patients, and drug overdose patients (Cooley, 1977; Gordon and Thompson, 1975). In intubated patients, the blockage of the air passage is also of major concern, since the vocal cords are rendered inoperative by the tube (Pope et al., 1974).

Methods of transduction

Respiration is very sensitive to oral pressure; therefore, the breathing instrument must function with very low flow impedance (Welkowitz and Deutsch, 1976). Respiratory monitors have been developed which use a wide variety of methods to detect respiratory activity and although most function acceptably well under certain types of conditions, none of the monitoring schemes have achieved high reliability under varying patient conditions (Cooley, 1977). Some of the schemes used are described as follows:

- Changes in temperature caused by airflow is detected by a thermistor positioned in the posterior pharynx via a nontoxic, flexible catheter with thin leads passing out through the nose (Gordon and Thompson, 1975; Jacobson and Webster, 1977). As the posterior pharynx receives both nasal and oral airflow, inhaling through either causes temperature to fall and hence changes the thermistor's resistance. The reliability can be enhanced by electronic signal conditioning and detection circuits that respond only to predetermined minimum changes in flow.

- The airflow can also be detected by a thermistor or thermocouple placed inside the nostril or near the mouth (Cooley, 1977; Das, 1981; Gordon and Thompson, 1975; Thomas, 1974). The advantages of this method include low noise and simplicity of operation, while its disadvantages include its being somewhat uncomfortable, subject to errors caused by coughing and sneezing, and susceptible to changes in breathing patterns and ambient drafts.
- Changes in chest circumference can be detected by mercury strain gauges or pneumatic expansion tubes (Cooley, 1977; Cromwell et al., 1980; Das, 1981; Gordon and Thompson, 1975; Thomas, 1974). When the flexible tube encircling the patient's chest is stretched by inhalation, it elongates and becomes thinner, with the measured variable changing

appropriately. Although this method has the advantages of low noise and simplicity of operation, it has the disadvantages of being dependent on body position and unreliable with diaphragm breathing or respiratory like motions. Body movements can also cause dimensional changes in the tube unrelated to lung volume changes, and hence, cause artifacts.

- Changes in the chest diameter can be detected by a high-frequency, or ultrasonic, transmitter and receiver system (Cooley, 1977). The transmitter attached to the chest, and the receiver beneath the mattress, detect respiration by the changes in the distance between the two. Advantages of this method include low noise and no patient leads, while its disadvantages include a dependency on body position and an unreliablity with diaphragm breathing or respiratory like motions.
- Body motion caused by respiration can be detected by a pressure transducer in a thin air mattress (Cooley, 1977; Gordon and Thompson, 1975). Although this method involves no patient contact, it is subject to artifacts, also responding to motions other than that of respiration.
- The airflow in an external airway can be detected

by an impeller, pneumotachygraph (differential pressure), hot wire, and ultrasound (Cooley, 1977; Jacobson and Webster, 1977). Although this method has the advantages of low noise and high accuracy, it requires an external airway which increases dead-space, and hence, increases the effort in breathing.

Respiratory events can be detected by any scheme that monitors volume changes during the respiratory cycle, with some methods that are difficult to calibrate, capable of detecting relative changes well. Among these is electrical impedance pneumography, which although not suitable for accurately measuring lung volume changes, is one of the most extensively used methods for respiration rate detection (Cooley, 1977; Cromwell et al., 1980; Das, 1981; Jacobson and Webster, 1977). Respiration produces cyclic alterations in the electrical characteristics of the tissues of the thorax, with the impedance changes due primarily to changes in the conductivity of the current path. As the chest expands during inspiration, both the lung-tissue impedance and the impedance of the thoracic wall itself increases, with the lung-tissue filling with air, and the chest wall becoming thinner as the circumference increases. Thus, the volume changes in the thorax are reflected as impedance changes, which can be detected by passing a small high-frequency current through electrodes.

The impedance measurement can be made using two- or four-electrode systems (Cromwell et al., 1980). In the two-electrode system, a constant current is forced through the tissue between two electrodes and the resulting voltage changes measured between these same electrodes. In the four-electrode system, the constant current is forced through the two outer, or current electrodes, and the voltage is measured between the two inner, or measurement, electrodes. The internal body resistance between the electrodes forms a physiological voltage divider, and the small current flowing through the measuring electrodes reduces the possibility of errors due to changes in the electrode resistance.

Since the quality of the electrode interface is practically the sole determinant of the reliability of the respiratory monitor, impedance monitors require moderate care in affixing the electrodes to the patient (Cooley, 1977). Electrode placement is also moderately critical for impedance respiratory monitors, with those devices which use two electrodes performing the best when the electrodes are placed on the midaxillary line, one on each side of the chest, at the fifth or sixth intercostal space (Cooley, 1977; Jacobson and Webster, 1977). However, if the patient is restless, it may be advantageous to move the electrodes to the front of the chest (Cooley, 1977). Some two-electrode devices also require a reference electrode which can be located somewhere between the two.

Although this method has fairly high accuracy, is relatively simple to operate, and causes no patient discomfort, it has a number of problems associated with its use. First, electrocautery devices, defibrillators, and diathermy units all represent a potential threat to the sensitive circuitry of the impedance monitors (Cooley, 1977). It is also likely that the monitor will fail to operate properly while these other devices are in use on the patient, or if the patient contacts a low impedance path to ground. Second, since the impedance pneumograph detects transthoracic impedance changes as the lung volume varies in respiration, false positive signals may occur on electrode movement or chest wall motion (Gordon and Thompson, 1975). Third, improperly adjusted electrical impedance monitors can mistake cardiac activity for respiration, as there is a small thoracic impedance change associated with each heart beat (Cooley, 1977). In infants, thoracic impedance changes may persist in apnea due to pulsatile blood flow (Gordon and Thompson, 1975).

The block diagram of an electrical impedance pneumograph includes the following components (Cooley, 1977):

- Oscillator and current source - Most impedance monitors employ a high-frequency current to sense impedance changes. The circuits are designed to provide the constant AC current to the patient, while making it possible to detect impedance changes directly by monitoring the voltage drop across the patient using the same electrodes. Although high-frequency current may be applied to the patient without harm, with currents as high as 1 mA acceptable at frequencies in excess of 10 kHz, most monitors use frequencies in the range of 50 to 100 kHz, and currents less than 500 uA rms (Cooley, 1977; Das, 1981; Thomas, 1974).
- AC amplifier - The resultant high-frequency potential which varies slightly in amplitude with each breath, must be amplified to levels large enough to be used by the rest of the instrumentation circuitry.
- Demodulator - Once it has been amplified, the potential across the patient must be demodulated to produce a signal which reflects the changes of impedance due to respiration.
- DC amplifier - The respiration waveform must then be amplified, as the output of the demodulator yields both the undesired baseline impedance, 100

to 1500 ohms, and the much smaller desired change, 0.2 to 10 ohms (Cooley, 1977; Thomas, 1974). Capacitive coupling with a time-constant long enough to pass the respiratory frequency, 0.1 to 2 Hz, is used to extract the desired waveform (Das, 1981).

- Respiratory pulse generator - To reliably count respiration rate, most monitors feed the waveform into a circuit which generates a single sharp pulse for each respiratory cycle, i.e., a Schmitt trigger input to a one-shot.
- Conversion from period to frequency - Conversion to a rate value can be accomplished using any of the techniques described above for use in determining heart rate.
- Apnea alarm - Some monitors include a built in alarm which is triggered if no respiration is detected within a certain time limit, i.e., 10 to 20 seconds (Das, 1981; Pope et al., 1974; Thomas, 1974).
- Electrode fault alarm - Poor electrode contacts can result in erroneously high or low counts on the impedance rate monitor, and thus, some indication of electrode contact status is desirable. Many devices are designed to provide indication that the

interface impedance has become unacceptably high, usually 4 to 10 times expected value (Cooley, 1977).

- ECG monitor - The ECG waveform can be provided from the same electrodes, with the proper filter separating it from the carrier wave.

The front panel of a respiration rate monitor usually includes the rate displayed on a meter, adjustable alarm limits, alarm indicator lamps, electrode status lamp, a lamp indicating each breath detected, and a sensitivity adjustment (Cooley, 1977; Thomas, 1974). The sensitivity adjustment is needed because of the wide physiological variations between patients, which results in signals representing normal respiration differing markedly in amplitude. The lamp indicating the detection of a breath is valuable when making such adjustments, and provides a continuous indication of proper function.

The determination of rate

One of the major problems of measuring the average rate of respiration is its extremely long inter-breath interval, which results in a meter having a great deal of ripple or a very slow response. To overcome this problem, Ben-Yaakov and Cohen (1979) suggest the use of the rate multiplication principle, whereby the incoming rate is multiplied by a constant factor to produce a train of pulses of a much higher frequency. This train of pulses can then be processed by an averaging rate meter to produce an acceptable analog output. Two integrator circuits are used in the rate multiplication circuit, with the first producing a voltage proportional to the breath-to-breath period, and the second used with a comparator, to trigger a pulser at a rate linearly related to the previously produced voltage.

Since knowing the actual rate of respiration is often not as important as being notified of its cessation, several apnea monitors have been designed (Das, 1981; Chess et al., 1976; Pope et al., 1974). In general, these monitors consist of an analog integrator with a DC voltage input, which generates a magnitude-limited output that increases linearly with time at a rate proportional to the input value. An audible alarm is enabled when the time interval between successive reset signals is such that the output of the integrator increases past a preset level. If, however, a reset signal is applied to the integrator before its output reaches this preset level, the integrator is reset to zero and the cycle is repeated.

The Monitoring of Temperature

As one of the oldest known indicators of the general well being of a person, the measurement of body temperature is considered to be one of the vital signs of medicine, being of great importance in diagnosis and treatment. Two basic types of temperature measurement can be obtained, systemic and skin surface, and although both provide valuable diagnostic information, the former is more commonly used. Systemic temperature of the internal regions of the body is maintained through a carefully controlled balance between the heat generated by the active tissue, mainly the muscles and the liver, and the heat lost to the environment (Cromwell et al., 1980). This control system is known to be affected by a group of substances, known collectively as pyrogens, in the bloodstream (Cromwell et al., 1980; Spooner, 1977b). An imbalance of the control system caused by such things as the presence of infection or the products of tissue destruction, permits the temperature to go higher than normal, as though the thermostat in the brain was turned up.

Many locations for core temperature measurement have been investigated, but none have had the widespread acceptance of systemic temperature measurements made in the mouth or the rectum, with the esophagus sometimes used in unconscious patients (Cromwell et al., 1980; Jacobson and Webster, 1977; Welkowitz and Deutsch, 1976). Although for hygenic reasons most temperature measurements are made in the mouth, oral temperature is about 0.5°C lower than rectal temperature, as it is somewhat dependent on the temperature of the environment (Jacobson and Webster, 1977). Thermometers which measure core temperature should be able to monitor the range of 35° to 44°C (Thomas, 1974; Welkowitz and Deutsch, 1976).

Except for the narrower range required and the differences in the size and shape of the sensing elements, instruments for the measurement of temperature in the body differ very little from those found in industrial applications. A high degree of accuracy is not always important, but the methods of measurement must be reliable and easy to perform, with the sensors fashioned for its own particular sensing location, small enough that response time is short and measurement localized (Cromwell et al., 1980; Welkowitz and Deutsch, 1976). An indicating type of electronic module is usually all that is required, needing only to be connected to the thermometer probe, and except for periodic zeroing and calibration, having no operating controls (Spooner, 1977b). As the sensor probe and cable used in continuous monitoring systems often causes discomfort to the patient, it is usually more suitable for use on an unconscious patient (Cromwell et al., 1980; Jacobson and Webster, 1977).

Methods of transduction

Where continuous monitoring is not required, mercury thermometers are still the standard method of measurement, as they are inexpensive, easy to use, and sufficiently accurate (Cromwell et al., 1980). Electronic thermometers have become available as replacements of the mercury thermometers, with the thermistor- and thermocouple-based instruments being the first to appear. In thermocouples, the junction of two dissimilar metals produces an output voltage nearly proportional to the temperature at that junction with respect to a reference junction (Cromwell et al., 1980). Although they are low in cost and reliable as temperature transducers, they have several disadvantages which generally make them difficult to use. These include a low output voltage, a reference junction which must be maintained at a known temperature, and an excitation current through the thermocouple circuit which must be minimized to avoid heating at the junction (Ben-Yaakov and Sanandagi, 1976; Cromwell et al., 1980). Additional errors can be caused by the Peltier effect, where one junction is warmed while the other is cooled. Ben-Yaakov and Sanandagi (1976) suggest a circuit which reduces the complex electronics and temperature control of past designs, while avoiding the above problems, by using inexpensive linear CMOS integrated circuit chips as low-level signal conditioners and a bucking voltage for reference junction compensation.

A thermistor is a thermally sensitive semiconductor whose material possesses a high negative temperature coefficient and whose resistance varies with the absolute temperature in an inverse exponential fashion (Cromwell et al., 1980; Thomas, 1974; Welkowitz and Deutsch, 1976):

$$Rt = Ro \exp\{k[(1/T) - (1/To)]\}.$$

Although its disadvantages include a nonlinear relationship between resistance and temperature, a danger of error due to self-heating, the possibility of hysteresis, and changing characteristics due to aging, its high sensitivity to small temperature changes and high stability made it universally accepted as the appropriate means for measuring patient temperature with medium to fairly high accuracy (Cromwell et al., 1980; Spooner, 1977b). The circuits can be linearized if high accuracy is required (Stockert and Nave, 1974), but the range over which thermistors are used to measure body temperature is relatively small and therefore in many cases the nonlinearity is unimportant (Spooner, 1977b). A typical thermistor probe temperature monitoring instrument is just a Wheatstone bridge circuit, with the probe forming one arm of the bridge, and the currents resulting from the bridge unbalance driving a current indication meter (Spooner, 1977b; Thomas, 1974). Dybvik (1976) and Rufer (1974), however, suggest schemes which use constant current bridge networks, instead of the constant voltage configuration, to avoid problems with unwanted voltage drops in the measuring bridges, thereby increasing linearity.

Electronic thermometer circuits that use semiconductor diodes or transistors as its sensors were the next to appear. The voltage across the diode or the base-to-emitter junction of a transistor changes linearly with temperature, varying at a rate of -2.2 mV per degree Celcius, with constant current excitation of the sensor ensuring that any voltage change across the junction is a direct result of the temperature changes at the probe (Koch, 1976; Nezer, 1977). A number of circuits have been suggested which take advantage of this temperature sensing scheme (Elmore, 1976; Koch, 1976; Nezer, 1977; Wurzburg and Hadley, 1978).

Semiconductor integrated circuits have been introduced to the market for use as temperature sensors. Current mode transducers, i.e., the LM134/334 by National Semiconductor Corp. (Cole, 1977) and the AD590 by Analog Devices Inc. (Electronics, 1977), produce output currents proportional to the absolute temperature when driven by a DC supply voltage. The range of these devices is a great deal less than that of more traditional temperature sensors, but they out-perform thermocouples and thermistors in terms of linearity. The LM134/334 requires a single external resistor for temperature sensing applications, needing to be trimmed at only one point for a 1% slope accuracy. The AD590, meanwhile, is laser trimmed to produce 298.2 uA at 25°C, with the output changing at a rate of 1 uA per degree Celcius. These current mode transducers are ideal in remote sensing applications, as series resistances in long wires do not affect their accuracy and only two wires are required. National Semiconductor (1977a) suggests a thermometer which uses the LM134, while Kraengel (1980) uses the AD590K. Voltage mode temperature transducers, such as the LX5700 (National Semiconductor) which produce 10 mV per degree Kelvin, are also available, and thermometers using them as the sensor probes have been suggested by Swift (1979) and Fox (1980).

The Monitoring of Arterial Blood Pressure

Blood pressure is considered to be a good indicator of the status of the cardiovascular system (Furno, 1981c). For normal body function, the pressure in various organs should lie within certain limits, with both an increase, hypertension, and a decrease, hypotension, being signs that usually indicate conditions calling for medical treatment (Jacobson and Webster, 1977). For a thorough hemodynamic examination of the cardiovascular system, a complete pressure waveform is required; the indirect method of measuring blood pressure using a sphygmomanometer is thus inadequate, and a direct method must be used in which the vessel is punctured (Jacobson and Webster, 1977; Welkowitz and Deutsch, 1976). Arterial pressure is defined as pressure where the site of measurement is distal to the aortic valves and proximal to the capillaries, and is taken as an indication of the performance of the left ventricle of the heart (Fox, 1977). It is also a good indicator of, and much more readily measurable than, cardiac output, and together with other data, allows significant parameters such as vessel compliance, blood volume, and peripheral resistance, to be inferred.

The heart pumping cycle and its associated pressure waveform is repetitive and can be divided into two major parts (Cromwell et al., 1980; Fox, 1977). Systole is the period of contraction of the heart muscle, specifically the ventricles, at which time blood is pumped into the pulmonary artery and the aorta, with systolic pressure, being in the range of 100 to 175 mm Hg (Fox, 1977; Jacobson and Webster, 1977; Thomas, 1974), defined as the greatest pressure reached during one cardiac cycle. Diastole, on the other hand, is the period of dilation of the heart cavities as they fill with blood, with diastolic pressure, being in the range of 60 to 100 mm Hg (Fox, 1977; Jacobson and Webster, 1977; Thomas, 1974), defined as the minimum pressure obtained in one cardiac cycle. Other parameters of interest include the pulse pressure, which is the difference between systolic and diastolic pressure, and the mean pressure, which is the true time-weighted average over one cardiac cycle, approximately diastolic plus one-third of pulse pressure (Fox, 1977).

Methods of transduction

In routine clinical tests, blood pressure is measured indirectly using a sphygmomanometer, which although easy to use, results in a somewhat subjective systolic and diastolic arterial pressure reading (Cromwell et al., 1980). Indirect pressure readings can be automated by such methods as oscillometry, acoustics, optics, doppler ultrasound, and impedance plethysmography, though continuous recordings have not been possible, as the practical repetition rate is limited (Cromwell et al., 1980; Jacobson and Webster, 1977). Of the methods that have been developed, all function satisfactorily in the patient with normal blood pressure, but most fail when the pressure is low, as in shock patients where the monitoring is most needed. However, this does not mean that the indirect method of pressure monitoring is not of great value in intensive care.

The direct measurement of blood pressure includes any determination of pressure wherein the measurement system comes into contact with the bloodstream. Although this method provides a continuous detailed pressure contour and is considerably more accurate than indirect methods for determining systolic and diastolic pressures, the patient's condition must warrant the potential risks involved with the invasion of the vascular system (Cromwell et al., 1980).

For the direct measurement of blood pressure, the pressure source, some point in the vascular system of the subject, must be coupled to the pressure sensing element of the transducer, which converts the mechanical energy of the applied pressure into electrical energy (Fox, 1977). This coupling of the pressure source with the transducer can be accomplished in several ways (Cromwell et al., 1980; Fox, 1977; Furno, 1981c):

- A catheterization method in which a liquid column in the catheter transmits the blood pressure from the sensing port at the catheter tip to the diaphragm of an external transducer.
- A catheterization method involving the placement of the transducer at the actual site of measurement in the bloodstream, either by passing the transducer through a catheter or by mounting it on the tip.
- A percutaneous method in which the blood pressure is sensed in the vessel just under the skin by the use of a needle or catheter.
- An implantation technique in which the transducer

is more permanently placed in the blood vessel or the heart by surgical methods.

As these methods of pressure measurement involve direct contact with the bloodstream, several precautions must be taken. First, all surfaces contacting the bloodstream or contacting, and including, the fluid column in the catheter must be sterilized (Fox, 1977). Second, although catheterization is relatively safe when experienced personnel use good technique, the catheter could damage the arterial wall if care is not exercised (Jacobson and Webster, 1977). Further, extrasystoles have been observed when the tip of the catheter touches the heart muscle. Finally, because the catheter invades the vessel, it is necessary to flush the tip every few minutes to prevent the clotting of blood and the formation of emboli (Furno, 1981c; Jacobson and Webster, 1977; Thomas, 1974).

With the pressure transducer connected to a patient's bloodstream by a conductive fluid, care must be taken that no stray electrical currents are conducted through the fluid column to the patient's heart. Since the catheter acts as a good insulator, all leakage current would pass to the tip, with currents as small as 10 uA having the capability of causing ventricular fibrillation (Jacobson and Webster, 1977). Therefore, precautions must be taken to insure that the patient is isolated from ground, so as to prevent current flow in the event of a transducer failure (Fox, 1977). The transducers should further be able to withstand the extreme voltages of defibrillation and electrocautery without failing.

A pressure transducer has two inputs, an excitation signal and an applied pressure, and outputs an electrical signal dependent on both (Fox, 1977). Although an ideal transducer would have a linear response for changes in either excitation voltage or applied pressure, many transducers exhibit a nonlinear response to changes in the excitation signal, which results from the heating effects of the currents involved. The linearity of response to applied pressure, however, is quite good, though in general, most transducers have a slight hysteresis. Also, although an ideal transducer would have an output signal of 0 V with no applied pressure, all transducers have some offset voltage which may vary with either time or temperature and can affect measurement accuracy appreciably once the signal conditioner has been adjusted.

Transmission problems are associated with the use of fluid-filled catheters in measuring blood pressure, with factors such as volume displacement, geometry, and mass of fluid, affecting the resonant frequency of the system (Fox, 1977; Jacobson and Webster, 1977). Most transducers are highly underdamped, and exhibit marked resonance in their frequency response (Fox, 1977). The fluid column used, also has a natural frequency or resonance of its own, that can affect the frequency response of the system (Cromwell et al., 1980). Thus, the transducer and catheter must be selected together, such that the upper limiting frequency of the complete system is adequate, with an optimally damped system, 0.7 damping factor, accurately reproducing an arterial pressure waveform (Cromwell et al., 1980; Fox, 1977; Jacobson and Webster, 1977). As catheter tip transducers have no fluid system between the pressure measuring site and the pressure sensing element, they have a higher frequency response and a more uniform frequency characteristic (Fox, 1977; Furno, 1981c; Jacobson and Webster, 1977). They are, however, expensive and prone to damage (Furno, 1981c). Fox (1977) reports that the harmonic analysis of the blood pressure waveform of humans shows that the amplitude of the 5th harmonic is 5 to 15% of the fundamental, while that of the 10th harmonic is 1.5 to 2% of the fundamental, and that of the 15th harmonic is 1% of the fundamental, thus making the required frequency response of the measuring system dependent on the fundamental frequency. Several sources (Cromwell et al., 1980; Fox, 1977; Jacobson and Webster, 1977) have recommended the use of a frequency range of DC to 30 Hz, with DC being necessary to measure the mean component, and 30 Hz to include the 15th harmonic at

120 beats per minute. Others (Furno, 1981a; Thomas, 1974; Welkowitz and Deutsch, 1976), however, have suggested that larger frequency ranges be used.

The most common type of pressure transducer is a mechano-variable resistance, or strain gauge, which depends upon the physical change in length or diameter of a wire resistance element due to stretching or other deformation (Cromwell et al., 1980; Fox, 1977; Thomas, 1974). The strain elements, usually connected as part of a Wheatstone bridge, are attached to the pressure diaphragm, bonded or unbonded, in such a way that as pressure increases, two stretch while two contract, with the changes in resistance unbalancing the bridge (Cromwell et al., 1980; Fox, 1977). Advantages of this type of transducer include small size, good temperature stability, high accuracy, low sensitivity to vibration, and the capability of being excited by either an AC or DC signal, while its primary disadvantage is its relatively low sensitivity (Fox, 1977).

In the differential transformer transducer, two secondary coils are wound oppositely and connected in series (Cromwell et al., 1980). The transducer detects changes in the magnetic coupling between these coils when the magnetically permeable core material, coupled to the elastic member, is moved relative to the coils, changing the symmetry, and producing a voltage change across the combined coils (Cromwell et al., 1980; Fox, 1977). Advantages of this type of transducer include high sensitivity and low hysteresis, while its disadvantages include sensitivity to vibration, low natural frequency, and a requirement of AC excitation, necessitating a demodulator in the signal conditioner (Fox, 1977).

In the capacitive transducer, one plate of a parallel plate capacitor is coupled to the elastic member (Cromwell et al., 1980; Fox, 1977). The transducer then detects the changes in the capacitance when the area of its plates or the distance between them is varied by the changes in applied pressure. The advantages of this type of transducer include good frequency response due to an extremely small displacement volume, low sensitivity to vibration, and high sensitivity to applied pressure, while its disadvantages include a high-impedance output, poor temperature stability, lead wires which introduce errors in capacitance, and a requirement of AC excitation (Cromwell et al., 1980; Fox, 1977).

The bonded silicon element bridge produces a sizable change in output voltage for a small displacement of the pressure sensing diaphragm, even at low DC voltage excitation (Cromwell et al., 1980). However, this high sensitivity is obtained at the expense of temperature stability. Delaunois (1974) reports obtaining highly satisfactory results in tests conducted on the LX1600A (National Semiconductor), an absolute pressure transducer, when it first became commercially available. Calibrations performed with the transducer in the physiological range showed perfect linearity, an output of 0.75 V per 100 mm Hg, and no detectable hysteresis. NSC has since developed a line of pressure transducers of varying configurations and specifications for different applications (National Semiconductor, 1977b; National Semiconductor, 1979a).

In a direct-coupled pressure amplifier system, a DC voltage acts as the excitation signal to the transducer, which then returns a DC signal proportional to the pressure input (Fox, 1977). As the frequency response of the amplifier is usually much higher than required by the signal or the transducer-catheter system, the returned signal is filtered to eliminate electromagnetic interference before further processing takes place (Fox, 1977). The balance control sets the output of the amplifier to 0 V when only atmospheric pressure is applied to the transducer, and is always necessary due to variations in atmospheric pressure and the physical positioning of the transducer (Fox, 1977; Thomas, 1974). The amplifier sensitivity control then sets the gain and calibrates the amplifier when some known pressure, at or near full-scale, is applied to the transducer.

The determination of pressure parameters

A useful aid in the interpretation of the changes that may occur in the amplitude of the pressure waveform is the separation of the systolic and diastolic peak levels, as well as the detection of the mean pressure level (Fox, 1977; Francis, 1974). These can then be used to drive various displays and indicate alarm conditions, with the front panel of an arterial pressure monitor generally including the systolic and diastolic/mean display, adjustable high/low alarms, alarm indicator lights, diastolic/mean selector switch, sensitivity adjust, and zero adjust (Thomas, 1974).

Mean pressure can be detected by severely restricting the bandwidth, with all fequency components except DC effectively removed. High-frequency cut-off is usually about 0.08 Hz, which corresponds to a time-constant of 2 seconds, though some time-constants as long as 10 seconds have been used (Fox, 1977). A compromise must be made, for although a longer time-constant reduces output ripple, the output responds more slowly to the changes in mean pressure.

Systolic/diastolic pressure separators are basically peak detectors, with the most important specifications being accuracy and time-constant (Fox, 1977). The choice of values for the peak detector capacitors is dependent on the acceptable fall rate of the voltage detected, with the value selected being a compromise between an output with a high ripple content and one that leaves the detector incapable of following rapidly changing waveforms. A selection of capacitor values is, therefore, usually provided. Two schemes for the detection of systolic and diastolic pressures have been described by Francis (1974) and Tjandrasa (1981).

In the scheme suggested by Francis (1974), the blood pressure waveform signal is applied via a 70 Hz low-pass filter to the input of a buffer amplifier. The signal is then applied to a positive peak detector (an "ideal diode" with a voltage follower) resulting in a voltage proportional to the systolic pressure. The signal is also applied to a precision clamp via a coupling capacitor, effectively grounding the most negative-going point of the pressure waveform, with the resulting peak-to-peak variations going to a second positive peak detector, which produces a voltage proportional to the pulse pressure. Once the pulse pressure and the systolic pressure have been derived, subtraction of the former from the latter yields a value proportional to the diastolic pressure.

The peak-and-valley detector described by Tjandrasa (1981) also uses two positive peak detectors. The output of the first positive peak detector gives the peak value of the input voltage waveform. Subtracting the input voltage from this peak value and applying the resulting waveform to a second positive peak detector yields the peak-to-valley value. The valley value can then be obtained by subtracting the output of the peak-to-valley peak detector from the output of the peak-input peak detector.

CIRCUIT DEVELOPMENT AND DESCRIPTION

As was seen above, heart rate monitoring can be easily combined with either the monitoring of arterial blood pressure or that of respiration rate, while temperature monitoring essentially stands alone. It was therefore decided that the parameters to be monitored would be divided into two units, with the first monitoring heart rate, respiration rate, and temperature, while the second monitored the blood pressure parameters and pulse rate. Below are the circuit descriptions of the complete heart rate/respiration rate/temperature monitor and the front-end of the arterial pressure/pulse rate monitor.

Selection of Integrated Circuits

Besides the proper functioning of the circuit at the supply voltages available, the primary concern when designing battery-operated instrumentation is that the circuit draws a minimal amount of current. In analog devices, there is often a tradeoff between the power supply requirements and the device characteristics, i.e., the input specifications of offset voltage and bias current, output specifications of source and sink currents, and the speed specifications of slew rate and bandwidths. In digital devices there is also a tradeoff between the power supply requirements and the device characteristics, i.e., noise margins, propagation delays, output drive currents, and clock speeds. In describing the integrated circuits selected, only those specifications which are of interest for this application will be mentioned. For more information, the interested reader should consult the National Semiconductor Linear, CMOS, and Special Function Databooks, the Intersil Data Book, and those application notes and specification sheets referenced below.

As operational amplifiers are the building blocks for most analog circuits, the two selected for use will be treated in some detail, with all other integrated circuits used described as they appear. The LM124 series of integrated circuits consist of four independent, high-gain, internally frequency compensated operational amplifiers (National Semiconductor, 1980). These op amps were designed specifically to operate from a single power supply over a wide range of voltages, though operation from split supplies is also possible, with the low current drain being independent of the supply magnitude. Even though operated from only a single power supply, the LM124 series of op amps, unlike other op amps, has an input common-mode voltage range which includes ground and an output voltage which can also swing to ground in the linear mode. However, when using a single supply, caution should be taken that the input not go negative more than -0.3 V. In cases where this may occur, protection by an input clamp diode with a resistor is suggested. Typical characteristics of the LM324 include (National Semiconductor, 1980):

DC voltage gain	100 dB
Unity gain bandwidth	1 MHz
Power supply range - single	3 to 30 V
- dual	±1.5 to ±15 V
Supply current drain	800 uA
Input bias current	45 nA
Output current - source	40 mA
- sink	20 mA

Although single-supply operation is convenient and low current drain desirable, the rather high input bias currents of the LM324 make it unacceptable for some of the circuits required. In designing the LM108 series of operational amplifiers, the primary objective was to obtain very low input currents without sacrificing offset voltage or drift, with a reduction of power consumption being of secondary concern (National Semiconductor, 1969b; National Semiconductor, 1980). Although the speed of the amplifier was of little concern and is not spectacular, it is usually not a critical problem in high-impedance circuitry. The characteristics of this device make possible many designs not practical with conventional amplifiers and is ideally suited for applications requiring low input currents such as sample-and-hold circuits, timers, and integrators, or those requiring reduced power consumption. Although this integrated circuit is interchangable with, and uses the same compensation as the LM101, an alternate compensation scheme can be used to make it particularly insensitive to power supply noise. Typical characteristics of the LM308 include (National Semiconductor, 1980):

Dual supply voltages	±2 to ±20 V
Input bias current	7 nA
Supply current	300 uA
Input resistance	40 Mohm

The Heart Rate/Respiration Rate/Temperature Monitor

The circuit diagram of the Heart Rate/Respiration Rate/Temperature monitor is shown in Figures 1 through 15, with the integrated circuits used, listed in Table 1. In the figures, all resistance and capacitance values are in ohms and farads respectively, with +V equal to +9 volts and -V equal to -9 volts. It should be noted that some of the discrepancies between the design specifications and the specifications mentioned above are due to the monitor's use of needle electrodes as opposed to standard surface electrodes. Features of this monitor include:

- Beat-to-beat heart rate
- Breath-to-breath respiration rate
- Rectal temperature
- 3-1/2 digit LCD display
- Two display modes, cycle and select, with LED indicating parameter displayed
- Audio and visual alarm indication, with audio disable

- Adjustable high/low rate alarms
- Acardia and apnea alarms
- High impedance alarm
- Beat and breath indicators
- Battery operation with low-battery indicator

TABLE 1. IC labels used in Figures 1 through 15 and their corresponding integrated circuits

LABEL	IC	LABEL	IC	LABEL	IC
1	LM324	12	LF398	23	CD4013
2	LF352	13	ICM7556	24	ICM7556
3	LM324	14	LM308	25	LM339
4	LM308	15	LF398	26	CD4069
5	LM308	16	CD4053	27	CD4015
6	LM308	17	LH0094	28	CD4015
7	LM324	18	LF398	29	LM339
8	ICM7556	19	LF398	30	LM324
9	CD4001	20	CD4052	31	CD4069
10	CD4066	21	CD4011	32	LM334
11	LM308	22	CD4011	33	PCIM176

<u>Excitation, transduction, and processing of waveforms for rate monitoring</u>

The circuit used to produce the high-frequency constant-current excitation signal required by the impedance pneumography method of monitoring respiration rate is shown in Figure 1, where IC1a is in a Wein bridge oscillator configuration and IC1b is a voltage-to-current converter, which drives a floating load impedance. The Wein bridge oscillator takes advantage of the fact that the phase of the voltage across the parallel branch of a series RC network in series with a parallel RC network is the same as the phase of the applied voltage across the two networks at one particular frequency (National Semiconductor, 1969a). When this network is used as a positive feedback element around an amplifier, oscillation occurs at the zero phase shift frequency. To prevent the amplifier from going into saturation, one zener diode or the other breaks down when the amplitude of the output voltage exceeds a specified level, thus shunting the 10 kohm resistor, reducing the amplifier gain (Coughlin and Driscoll, 1977). The 25 kohm potentiometer allows for the adjustment of the output peak-to-peak value. As shown, the oscillator circuit is designed for an output frequency of 10 kHz and an amplitude of 8 V peak-to-peak, or 2.83 V rms. A higher frequency sine wave was not possible, for as frequency was increased, the amount of distortion also increased, possibly due to exceeding the LM324 slew rate.

Coughlin and Driscoll (1977) suggest that for best results, the oscillator's output be connected to a voltage follower to avoid undue loading of the circuit. However,

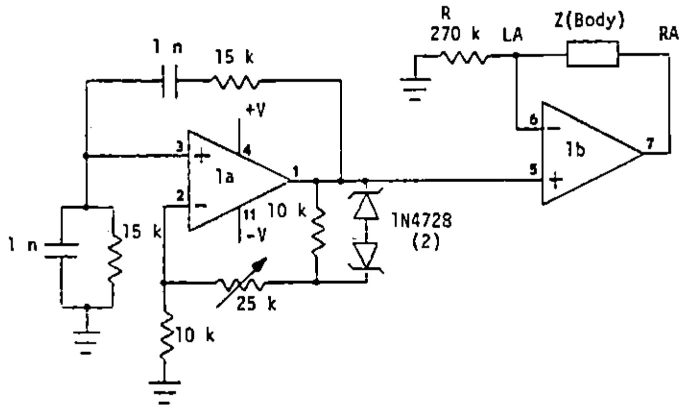


FIGURE 1. High-frequency constant-current excitation circuit

connecting the Wein bridge oscillator to the positive input terminal of the voltage-to-current converter results in the same high input impedance being seen by the oscillator; thus, no follower is used. Because of the virtual short at the op amp input, IC1b converts the Wein bridge oscillator voltage to a current equal to Vin/R, independent of the magnitude of the ungrounded load (Millman and Halkias, 1972). With R = 270 kohm, and Vin = 2.83 V rms, the current through the load impedance will be 10.5 uA rms. This current can now be introduced subdermally across the thorax of the subject using needle electrodes.

To sense the voltage changes of the high-frequency waveform across the thorax, as well as the ECG waveform, the instrumentation amplifier circuit of Figure 2 is used. The

LF352 is a monolithic JFET input differential current feedback amplifier, with the matched high-voltage JFET input devices providing very high input impedance and extremely low bias currents (National Semiconductor, 1980). Operating with an internal closed-loop gain connection which allows good linearity with no external feedback, these devices accurately amplify the differential-input signal while rejecting common-mode signals and noise. The gain, which is equal to Rs/Rg, is easily adjusted from 1 to 1000 by changing a single external resistor. Typical characteristics of the device, at gain = 100, include (National Semiconductor, 1980):

Input impedance	2×10^{12} ohm
Input bias currents	3 pA
Gain nonlinearity	0.02%
Common mode rejection ratio	120 dB
Supply current	1.2 mA
Small signal bandwidth	30 kHz
Slew rate	1 V/us

In the design shown, the gain of the amplifier is set at 100, with an output offset adjust provided, which at this gain is able to null the independent offset voltage contributions of the two amplifier stages.

Since the use of defibrillators or electrosurgical equipment can be destructive to the instrumentation amplifier, two protective measures are taken. Input stage failure of an IC op amp can be induced in two general ways, by exceeding the differential-input rating of the device, or by exceeding the common-mode rating (Jung, 1980). To protect the inputs from excessive differential voltages, a differential-input overvoltage/overcurrent circuit is used, in which an input clamping diode network limits the input differential voltage, while some series resistance is provided to limit the current during clamping. As the monitor is floating with respect to the defibrillator or electrosurgical device, any changes resulting from the use of these devices will cause the RL reference lead to shift along with it, thus preventing failure due to the destructive input current flow caused by exceeding the input common-mode range. In anticipation that defibrillator or electrosurgical equipment may cause the output of the amplifier to be abruptly changed more than 6 V, a PNP transistor is connected to prevent the slew rate of the output from exceeding the slew rate of the sense stage, thus avoiding degradation of the sense stage input transistor.

As the output of the instrumentation amplifier includes both the ECG waveform and the changes in thoracic impedance modulated by the carrier wave, it must be further processed before being used to trigger the inter-event interval detectors. The circuit used to separate the waveforms and demodulate the impedance changes is shown in Figure 3. Bandpass filters are designed to pass only a band of frequencies, having a maximum voltage gain, Ar, at the

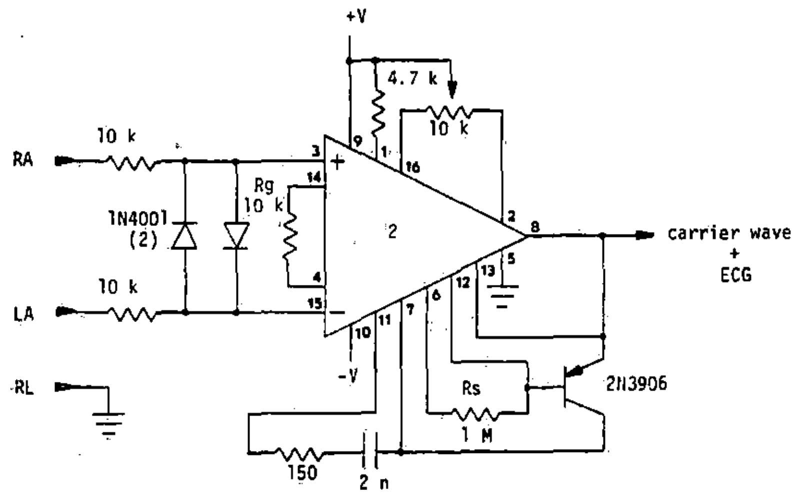


FIGURE 2. Instrumentation amplifier circuit with input and slew rate protection

resonant frequency, wr, while attenuating all other frequencies as they vary from resonance. As the waveforms of interest occupy two distinct regions of frequency, two bandpass filters are used to separate the instrumentation amplifier output into the desired components. One of the bandpass filters is used to pass the R-wave component of the ECG, while attenuating the P- and T-waves as well as the carrier wave, and the other is used to pass the carrier wave and its variations, while rejecting the ECG waveform. Using the design equations for active RC bandpass filters described by Millman and Halkias (1972), and Coughlin and Driscoll (1977), filters with the following characteristics were designed:

- R-wave bandpass filter, IC3a, as suggested by Thakor (1981)

Quality factor	3.3
Gain at resonance	1
Resonant frequency	17 Hz
Bandwidth	5.15 Hz

- Carrier frequency filter, IC1c

Quality factor	5.0
Gain at resonance	1
Resonant frequency	10 kHz
Bandwidth	2.0 kHz

To demodulate the changes in thoracic impedance, IC4 and IC5 act as a buffered peak detector which measures the maximum value of the fluctuating voltage. The circuit operates like an ideal diode with a storage capacitor, following the rising signal slope as the input exceeds the output, and holding its charged state when the input signal reverses its slope. To attain the necessary speed using the LM308, resistor R is included to allow IC4 to be clamped in the off-state by the clamping diode D, thus resulting in a faster recovery (Jung, 1980). The only means of capacitor discharge, in the absence of any bleed resistor, is the bias current of the buffer stage and the diode leakage.

To properly trigger the inter-event interval detectors,

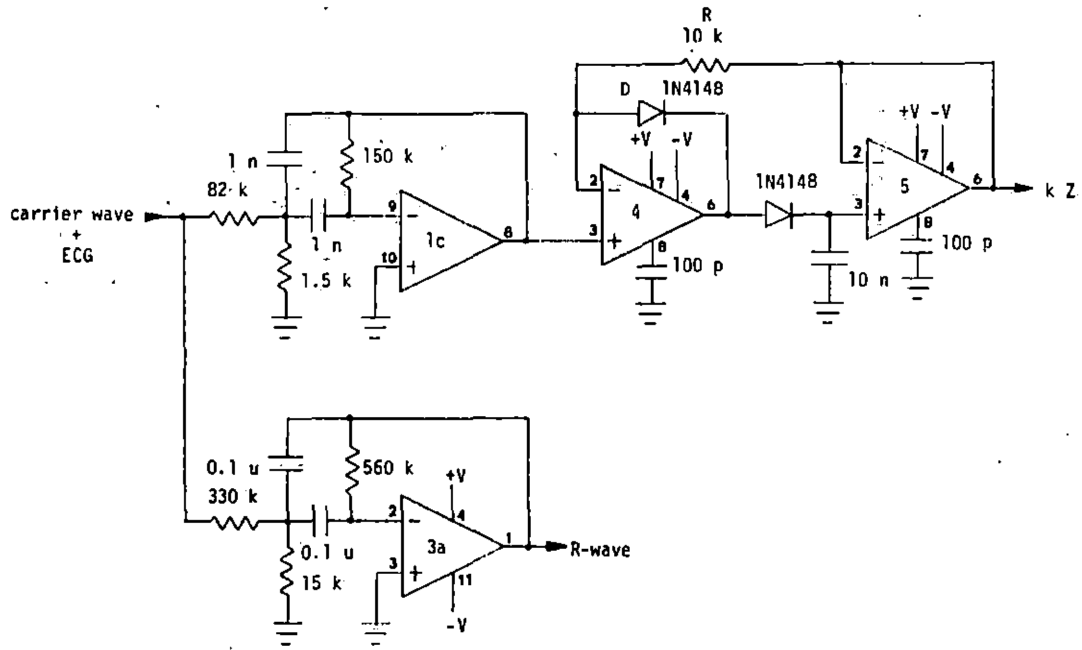
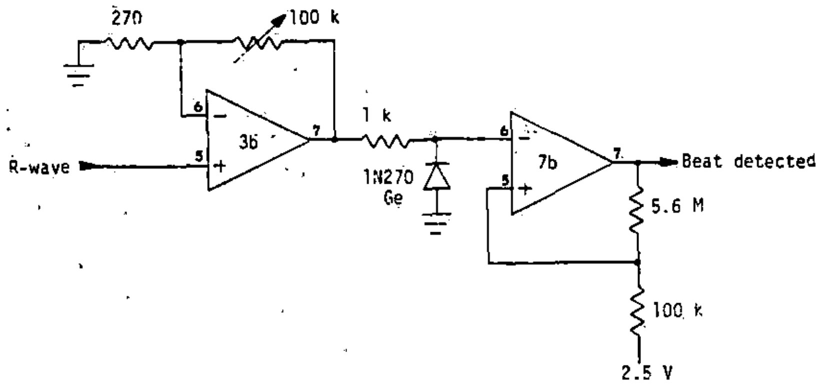
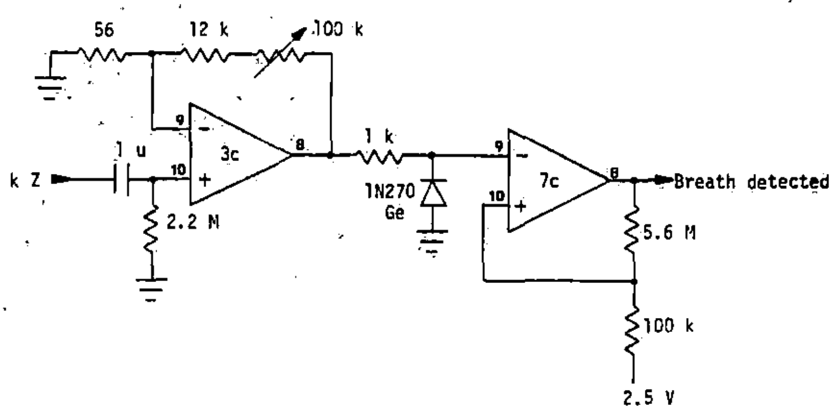


FIGURE 3. Waveform separation and impedance change demodulation circuit

each of the separated waveforms must be further processed to produce clean negative-going edges. Figure 4a shows the noninverting amplifier, IC3b, and the Schmitt trigger, IC7b, used to obtain the necessary triggering edge of the inter-beat interval detector. If the output voltage desired to guarantee a negative-going edge of the Schmitt trigger is 2.7 V, the input voltage of the Lead I R-wave is 0.07 to 1.13 mV, and the gain of the instrumentation amplifier is 100, then the gain of the noninverting amplifier must range between 24 and 386. As designed, the noninverting amplifier allows gains ranging from 1 to 371, i.e., from 0 to 51.4 dB.



(a)



(b)

FIGURE 4. Amplification and triggering on heart beat (a) and respiration (b)

The input common-mode voltage of the LM324 must be restricted for negative voltage swings, as an overvoltage which goes below -V (or ground) causes an internal diode to conduct. To prevent possible degradation, a clamp diode with the anode connected to -V (or ground) and a series resistor is used to restrict the input voltage (Jung, 1980). As IC3 has +V and -V power supplies, and IC7 has +V and ground supplies, the recommended circuit is used to prevent the negative input terminal of the Schmitt trigger from going more negative than the germanium diode cutin voltage, thus protecting the device.

The Schmitt trigger is a regenerative comparator, with the input voltage applied to the inverting terminal and the feedback applied to the noninverting terminal (Millman and Halkias, 1972; Coughlin and Driscoll, 1977). Positive feedback ordinarily leads to instability in an amplifier, but when used on a comparator, induces a snap action in the switching of the output from one limit to the other. The most important use of a Schmitt trigger is to convert a slowly varying input voltage into an output having an abrupt waveform, with the output transitions occurring at a precise value of input voltage. Although the inherent hysteresis of the circuit eliminates false output transitions if the threshold voltages are larger than the peak noise voltages, the input signal excursions must exceed the limits of the hysteresis range for proper operation. When the input signal exceeds the upper limit while the output is high, a negative transition results; conversely, exceeding the lower limit while the output is low, results in a positive transition. Using the circuit analysis of Millman and

Halkias (1972), the Schmitt trigger designed has a hysteresis of 0.16 V, an upper threshold of 2.61 V, and a lower threshold of 2.45 V.

A similar circuit is used to obtain the negative-going triggering edge of the inter-breath interval detector, though unlike the above circuit, a passive high-pass filter is required at the input to attenuate the large DC baseline impedance. As the -3 dB frequency of the high-pass filter corresponds to 1/[2(pi)RC], a breakpoint of 0.1 Hz (6 breaths/minute) requires that the RC time-constant be 1.59 seconds. As designed, with C = 1 uF and R = 2.2 Mohm, the time-constant is 2.2 seconds, corresponding to a breakpoint frequency of 0.072 Hz or 4.34 breaths/minute. If the output voltage of the noninverting amplifier desired to guarantee a negative-going edge of the Schmitt trigger is again 2.7 V, the range of impedance changes to be sensed is 0.2 to 10 ohms, the peak of the excitation current is 14.1 uA, and the gain of the instrumentation amplifier is 100, then the gain of the noninverting amplifier as calculated by:

$$
\begin{aligned}
G(NA) &= V(ST) \;/\; (delV)(G(IA)) \\
&= V(ST) \;/\; (Ipeak)(delZ)(G(IA))
\end{aligned}
$$

must range between 9574 and 191 (79.6 and 45.6 dB). If we relax the impedance sensitivity to 1 ohm, the required gain of the noninverting amplifier is 1915 (65.6 dB) giving a more reasonable range. With Rpotentiometer = 100 kohm, Rconstant = 12 kohm, and Ri = 56 ohm, the gain ranges from 215 to 2001, which corresponds to impedance changes of 8.91 and 0.957 ohms respectively. Thus, any impedance change greater than 0.957 ohm can be made to trigger the inter-breath period detector. By making the gain of the carrier wave bandpass filter 5, changes in impedance down to 0.2 ohm can be sensed.

Rate determination

The inter-event interval detector is designed to monitor frequencies ranging from 5 to 200 events per minute, with the lower limit reflecting the lowest rate of respiration in laboratory animals and the upper limit reflecting a hardware tradeoff. As it was desirable to have a fixed decimal position for design simplicity, while still being able to display temperature to the 0.1°C, the maximum value that could be displayed by the 3-1/2 digit panel meter selected was 199.9, thus the upper limit. This limit in no way reduces the usefulness of the monitor, as the animals for which it is intended - dogs and cats - have heart rates within this range.

When the physiological waveform exceeds the threshold of the Schmitt trigger, a negative-going pulse is produced, with the leading edge triggering the first monostable multivibrator. It is during this portion of the dead-time that the integrator value, a voltage proportional to the inter-event interval, is held constant and stored by the sample-and-hold circuit. The falling-edge of the first monostable multivibrator then triggers a second monostable, producing the remainder of the dead-time, during which the integrator is reset. The falling-edge of the second monostable then starts the integrate period of the cycle, and the process is repeated. The circuits used to determine the inter-beat and inter-breath intervals are shown in Figures 5 and 6 respectively, and their associated timing diagram is seen in Figure 7.

The monostable multivibrators (IC8a, 8b, 13a, and 13b) are designed using half of a ICM7556 dual general-purpose CMOS RC timer which provides significantly improved performance, i.e., typical supply currents of 160 uA, over the standard 556, while being direct replacements for those devices (Intersil, 1979). In a monostable multivibrator, a negative-going trigger pulse causes the output to switch from low to high, remaining high for a period determined by the resistor and capacitor connected to the IC. If the voltage at trigger terminal is held low, the output remains in the high state; thus, the width of the negative-going trigger pulse must be less than the width of the expected output pulse. In cases where the trigger terminal is to be held low, a resistor, capacitor, and diode pulse network

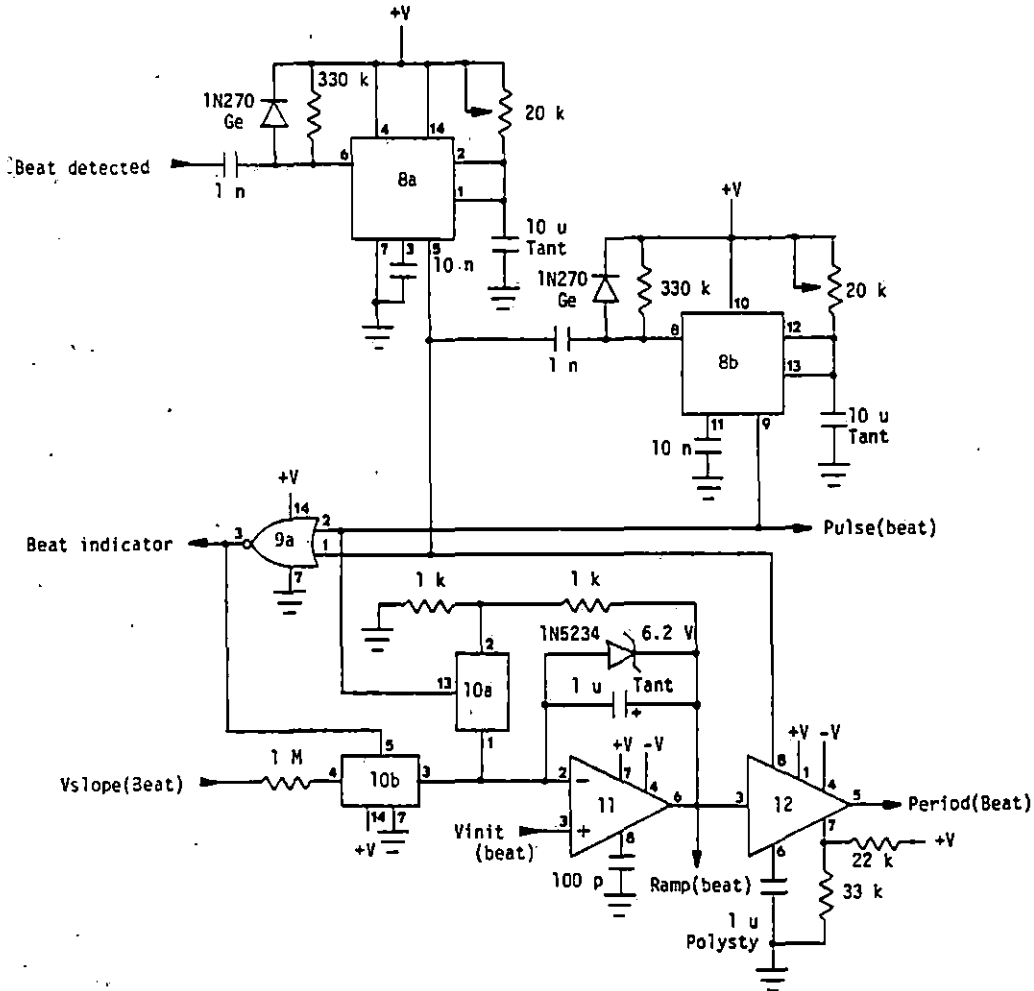


FIGURE 5. Circuit to determine the inter-beat interval

must be added to the input to ensure proper operation, with the RC time-constant being small with respect to the output timing interval (Coughlin and Driscoll, 1977). The resistor connected between Vcc and the trigger input ensures that the output remains low, and that the capacitor is charged to Vcc

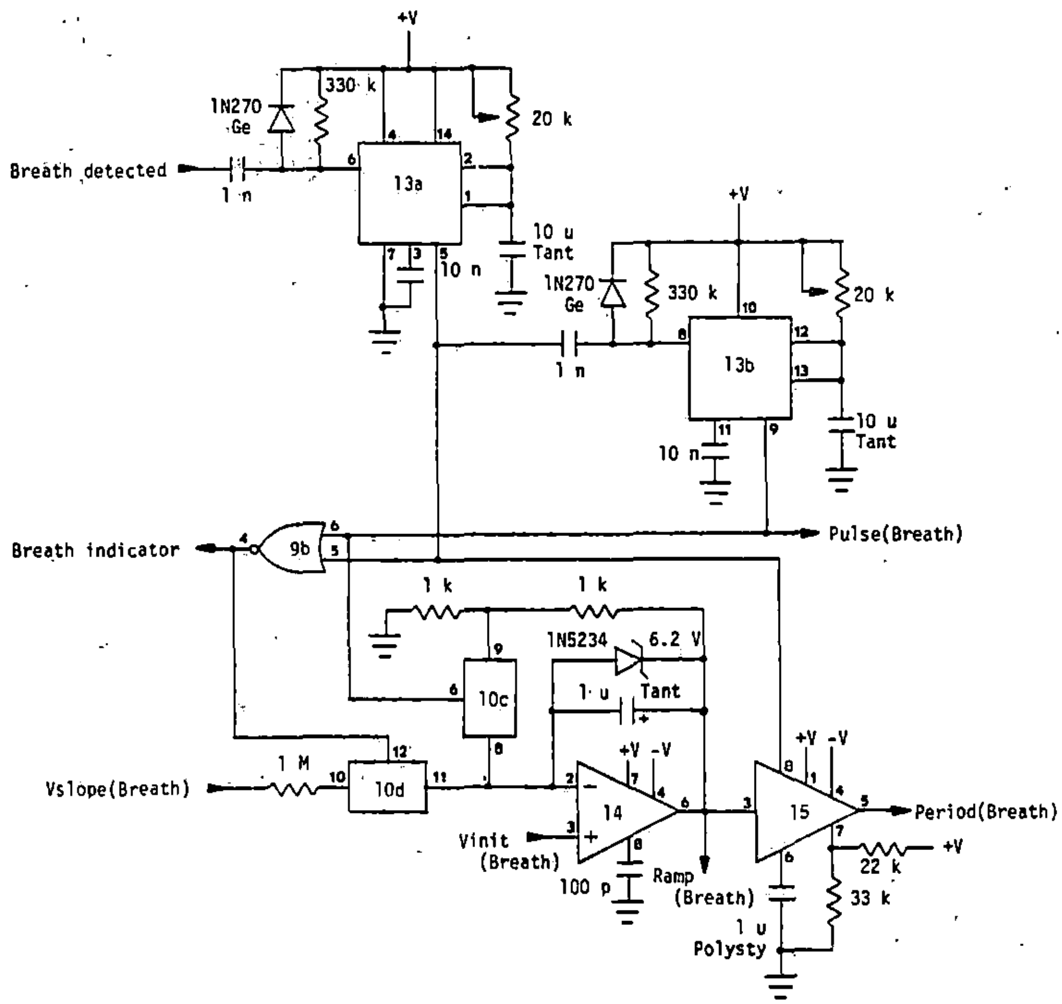


FIGURE 6. Circuit to determine the inter-breath interval

until a negative trigger pulse occurs. The germanium diode, meanwhile, prevents the timer from triggering on the positive-going edges of the input pulses, and also prevents the input from exceeding the supply voltage by more than the diode cutin voltage. With a maximum rate of 200

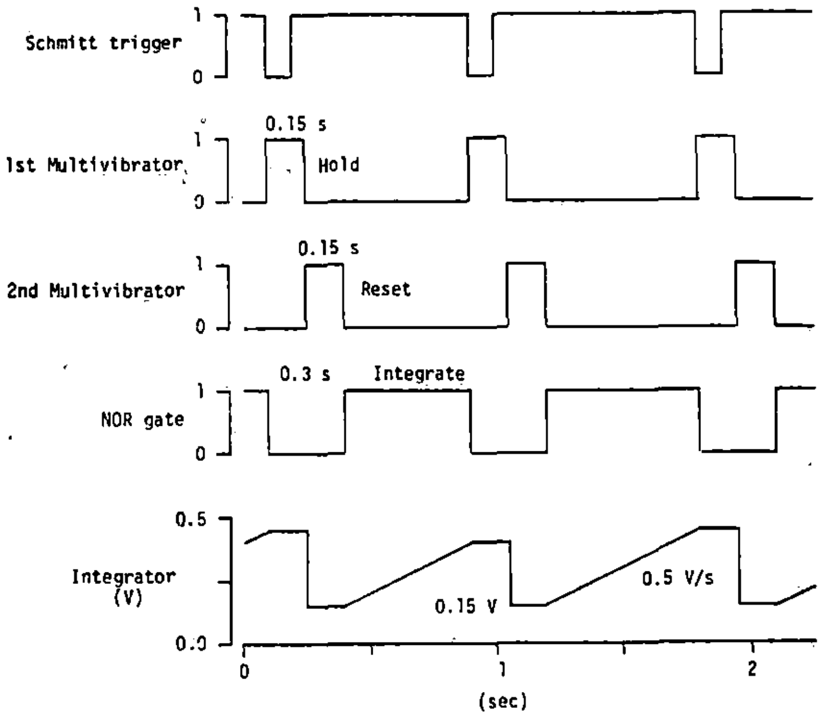


FIGURE 7. Timing diagram for the inter-event interval determination circuits

events/minute, corresponding to a dead-time of 0.3 seconds, the periods of the monostables were selected to be 0.15 seconds or 1/2 of the dead-time. For a capacitance of 10 uF, the required resistance would be 13.6 kohm, but to allow for capacitor tolerance, a 20 kohm potentiometer is used. The time-constant of the input pulse network is 330 usec.

The integrator circuits centered around IC11 and IC14, have reset capabilities and a magnitude limiter, performing the mathematical operation of integration to measure elapsed time. This circuit takes advantage of the fact that a constant current applied to a capacitor causes its voltage to change linearly at a rate $dV/dt = I/C$, and the fact that an op amp simulates an ideal current source (Jung, 1980). Since integration begins with whatever voltage is initially on the capacitor, the capacitor voltage must be reset to the desired value before each integration, sequentially cycling through the defined operating modes of reset (initialize), integrate, and hold, with the necessary switching directed externally by the pulses produced by the two monostable multivibrators which control the analog switches, IC10a to 10d.

The CD4066 integrated circuits are quad bilateral switches intended for the transmission or multiplexing of analog or digital signals. Typical characteristics include (National Semiconductor, 1978):

ON resistance	120 ohm
flat over the peak-to-peak signal range	
Variation between switch ON resistances	10 ohm
Distortion	< 0.4%
OFF switch leakage	0.1 nA
Quiescent device current	0.01 uA

Switch IC10b (10d) is open-circuited when the first multivibrator is triggered and remains open if either of the monostable multivibrators is in the high state, disconnecting the constant-current input to the integrator. During the first multivibrator period, switch IC10a (10c) is also open-circuited; thus, the output of the integrator is held at a voltage equal to the charge on the capacitor, plus the voltage at the positive input terminal. When the second multivibrator is triggered, switch IC10a (10c) closes, causing the integrator circuit to become effectively a noninverting amplifier with a gain of two. This forces the capacitor voltage to equal that at the positive input terminal, thus initializing the circuit.

The equations governing the design of the integrator circuit include

$$dVout/dt = -(I/C) = -(Vin/RC)$$

where the values of the resistor and capacitor determine the slope of the integrator, and the minus sign results from the inverting amplifier configuration used. Rearranging the above equation and integrating both sides yields,

$$\begin{aligned} dVout &= -(Vin/RC)\ dt \\ Vout &= -(1/RC) \int_{0.3}^{t} Vin\ dt + Vreset \\ &= -(Vin/RC)(t - 0.3) + Vreset \end{aligned}$$

for Vin constant. Letting the slope of the ramp, -(Vin/RC), be 0.5 V/s, defines

$$Vreset = (\text{dead-time})(Vramp) = 0.15\ V.$$

If R = 1 Mohm and C = 1 uF, then Vin = -0.5 V. Since the gain of the noninverting amplifier is 2, Vinit = Vreset/2 = 0.075 V

and since both inputs have an effect on the slope, by superposition

Vslope = Vin + Vinit = -0.425 V.

Both the Vinit and the Vslope reference voltages are designed such that they can be adjusted to take into account the resistor and capacitor tolerances, as well as the LM308 input currents and offset voltages. The maximum output of the integrator, as constrained by the 1N5234 zener diode, is 6.2 V which corresponds to 4.8 events/minute.

The high output of the first monostable multivibrator which causes the integrator output to be held constant, is also used as the logic input to the sample-and-hold circuit, IC12 (15), causing it to enter its sampling mode. The circuits used are of the LF198 series, which are monolithic sample-and-holds utilizing BI-FET technology to obtain ultra-high DC accuracy with fast acquisition of signal and low droop rate. Droop rates as low as 5 mV/min when using a 1 uF hold capacitor are reported (National Semiconductor, 1980). Typical characteristics of the LF398 include (National Semiconductor, 1980):

Supply voltage	±5 to ±18 V
Supply current	4.5 mA
Gain error	0.004%
Hold capacitor charging current for Vin - Vout = 2 V	5 mA

Leakage current into hold capacitor	30 pA
Hold-step (Ch = 1 uF)	0.01 mV

The high currents involved for both supply and charging are potential drawbacks, and require tradeoff considerations with the desirable device characteristics.

Although hold-step, acquisition time, and droop rates are usually major tradeoffs in the selection of hold capacitor values, the long acquisition time available (0.15 sec) allows for greater freedom. Further, the use of capacitors with low hysteresis such as polystyrene, polypropylene, or Teflon is suggested, as a significant source of error in an accurate sample-and-hold circuit is caused by the dielectric absorption of the hold capacitor (National Semiconductor, 1980). 1 uF polystyrene capacitors were used, though much smaller values would have served as well, i.e., 0.04 uF for Period(Breath), 0.05 uF for Period(Beat), and 0.06 uF for rates, as determined by the maximum potential hold period and the maximum allowable droop.

The decision to multiplex the digital panel meter also allows for the time-sharing of the analog divider integrated circuit, one of the more expensive ICs used. Figure 8 shows the circuit used to convert the input voltages proportional to the inter-event interval, into output voltages proportional to the corresponding frequency, for both heart rate and respiration rate. Two-thirds of a CD4053, IC16, triple two-channel digitally controlled analog switch IC is used to multiplex the divider input and demultiplex its output (National Semiconductor, 1978). This IC has three separate digital control inputs A, B, and C, which selects one of a pair of channels connected in a single-pole-double-throw configuration. Typical characteristics of this IC are similar to the CD4066 and include (National Semiconductor, 1978):

ON resistance	120 ohms
OFF channel leakage	0.01 nA
Variation in switch characteristics	10 ohms
Quiescent device current	40 uA
Distortion	0.04%

The LH0094, IC17, is a multifunction converter which generates an output voltage for the transfer function

$$Eo = y((z/x)^{**}m) \qquad \text{for } 0.01 \le m \le 10$$

and as such, can be configured for use as a precision divider or multiplier, as an exponential circuit, i.e., square and square-root, or as a logarithmic amplifier (National Semiconductor, 1979b). For basic operation, it is possible to use the device without any external adjustments or components, though where high accuracy is desirable, trimming of the gain and offset can be used to cancel all errors of conversion. As it is designed for positive input signals, a clamp diode is recommended for those applications in which the inputs may be subjected to open-circuits or

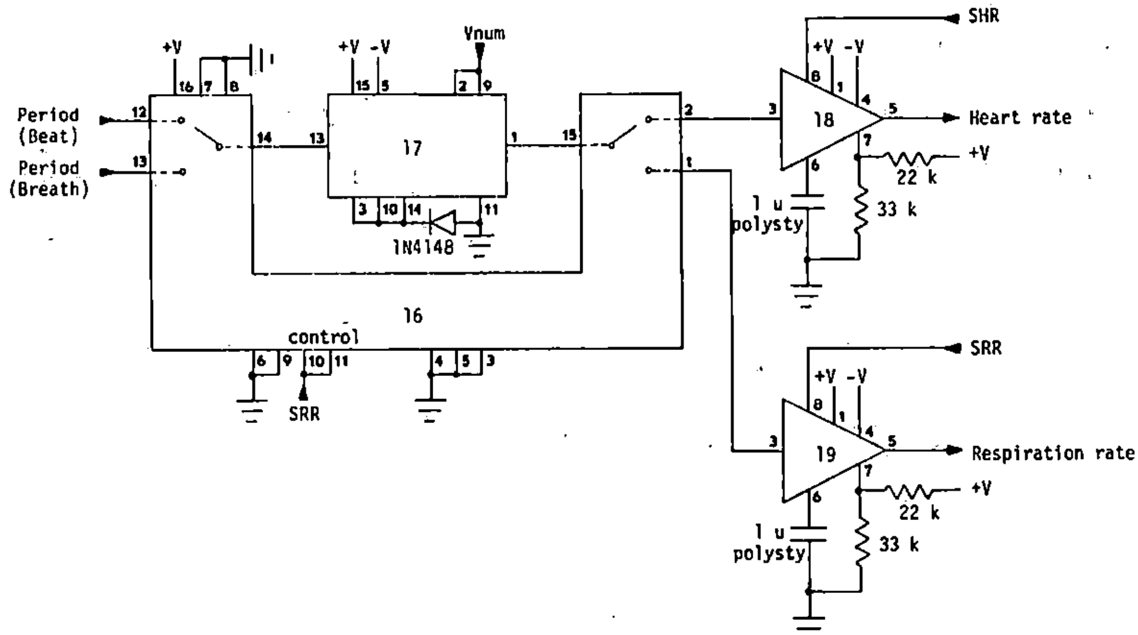


FIGURE 8. Circuit to time-share the analog divider

negative input signals. Characteristics of the analog divider include (National Semiconductor, 1979b):

Input impedance		100 kohm
Input voltage		0 to 10 V
Supply current		3 mA
Dual supply voltage		± 5 to ± 22 V
Accuracy of division		
Eo = 10 z/x at Vs = ± 15 and 0.01 ≤ z ≤ 10		
Untrimmed	0.5 ≤ x ≤ 10	0.45% FS
Trimmed	0.1 ≤ x ≤ 10	0.10% FS

As the output representation of the divider must equal 2.00 V for an inter-event interval of 0.3 seconds, i.e., a rate of 200 events/minute, and the input voltage is equal to (0.5 V/s)(T), the divider equation takes the form:

$$Vout = k/Vin = 0.3\ V^2/(0.5\ V/s)(T).$$

For the preliminary circuit, an untrimmed divider is used, with y = z = sqrt(0.3) = Vnum.

When employing a digital technique for the display of physiological variables, it is in many cases undesirable to have the beat-by-beat value displayed, since even in the stable patient, variations in the monitored parameter makes the value displayed difficult to read. To overcome this problem, Klevenhagen and Storey (1978) use a circuit which ignores parameter variations of a small but adjustable magnitude, using the DPM hold facility to retain the value on the display until the parameter exceeds its insensitive window. Berlin (1977) uses a 555, in an astable multivibrator configuration at the DPM external hold input, to override the internally controlled sampling period, thereby increasing the display time. As the digital display of our monitor is multiplexed, a simple solution to the flickering display problem is the use of sample-and-hold circuits to maintain the input of the DPM for the duration of the parameter's display period. No provision is made for holding the displayed value in the select mode.

Temperature transduction and processing

The third parameter to be monitored, body temperature, is sensed using a LM334, IC32, three-terminal adjustable floating current source, and calibrated using a summing amplifier IC7a (Figure 9). As mentioned above, these ICs are ideal for remote temperature sensing applications as current-mode operation makes accuracy independent of the series resistance of long wire runs, with only two lead wires required. The simplest one external resistor connection generates an output current directly proportional to the absolute temperature in degrees Kelvin according to the following equation (National Semiconductor, 1980):

$$Iset = (227\ uV/°K)(T) / Rset.$$

If Rset = 230 ohm and Rload = 10 kohm, then Vout = (10 mV/°K)(T).

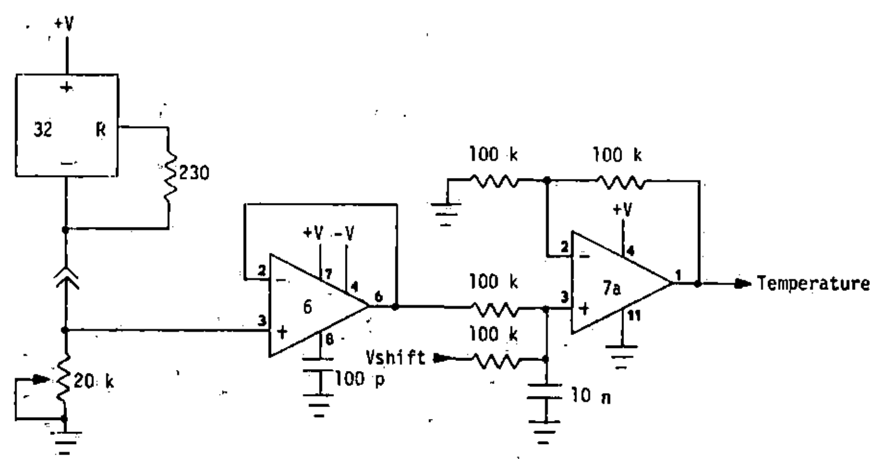


FIGURE 9. Temperature transduction and calibration circuit

To convert from a voltage proportional to temperature in degrees Kelvin to one that is proportional to temperature in degrees Centigrade, the buffered load resistor voltage is used as one of the inputs to a summing amplifier. This circuit, which gives an output equal to the algebraic sum of the inputs, is used to shift the current source load resistor voltage downward by 2.73 V to satisfy the equation:

$$°K = °C + 273.$$

Calibration of the temperature transducer is simplified because of the fact that most of the initial inaccuracy is due to a gain term (slope error) and not an offset (National Semiconductor, 1980). Therefore, a calibration consisting of a one-point gain adjustment will be sufficient to trim both slope and zero at the same time, because the output extrapolates to zero at 0°K independent of Rset or any initial inaccuracy. The necessary gain trim can be done on either Rset or on the load resistor used to terminate the current source, with the slope error after trim normally less than ±1%. To maintain accuracy, thermocouple and lead resistance effects should be minimized by locating the current setting resistor physically close to the device, while maintaining a constant Rset temperature if possible.

Digital display and associated logic

With all parameters having been properly scaled, the digital panel meter multiplexing scheme, shown in Figure 10, is discussed. Half of an ICM7556, IC24a, in the astable multivibrator, free-running, configuration will serve as the display system clock, with its output high and low state periods determined by an externally connected resistor-capacitor network (Coughlin and Driscoll, 1977). As shown, with C = 1 uF and RA = RB = 1 Mohm, the clock period is 2.09 seconds. Enabling or disabling the system clock will determine the mode of the display panel operation, with an enabled clock causing all parameters to be displayed in a prescribed cycle, and a disabled clock causing a selected parameter to be displayed continously. The mode selection is controlled by the select/cycle switch which is debounced by a S'-R' latch, IC21a and 21b, where L = 1 in the cycle mode and conversely, L = 0 in the select mode. For the clock to be enabled, the latch input to the NOR gate must be a logic 0. Thus, since the clock must be enabled when the switch is in the cycle position, L' is used as the NOR gate input.

IC23a and 23b, CD4013, are two D-flip flops in the Johnson counter configuration, where only one of the outputs is allowed to change at a clock transition. The BA sequence of the counter - 00, 01, 11, 10 - is used as the inputs to

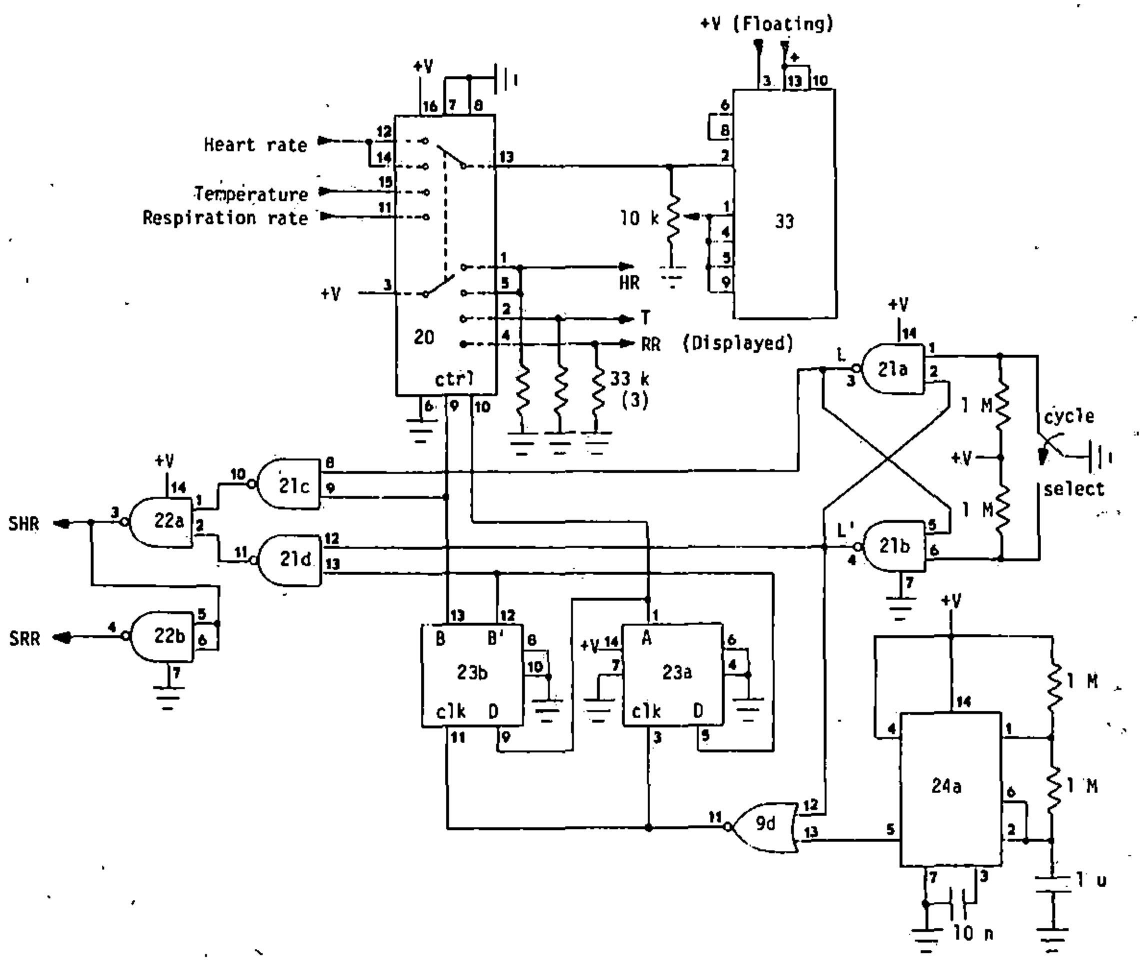


FIGURE 10. Digital display multiplexing and parameter selection circuit

the analog switch and the sample-and-hold logic circuit to control the sequence in which the parameters are displayed, i.e., heart rate, heart rate, respiration rate, and temperature respectively. IC20, a CD4052 dual four-channel digitally controlled analog multiplexer/demultiplexer switch, having the same specifications as the CD4053 above, is used to control the parameter being displayed (National Semiconductor, 1978). This multiplexer has two binary control inputs A and B, which select one of four pairs of channels to be turned on, with the upper channel used to transmit the appropriate input voltage to the DPM, and the lower channel used to transmit a logic 1 (9 V) to the LED display corresponding to the parameter being transmitted. The 33 kohm resistors at the outputs of the lower channel act as pull-down resistors when the output of the switch is open-circuited.

To update the heart rate and respiration rate values prior to being displayed, a sample-and-hold control/divider multiplexing logic circuit must be included (IC21c, 21d, 22a, and 22b). In the cycle mode, heart rate should be divided and stored while B = 1, i.e., when respiration rate or temperature is being displayed. Similarly, respiration rate should be divided and stored while B = 0, i.e., when heart rate is being displayed. In the select mode, on the other hand, heart rate should be divided and passed through when B = 0, while the same should be done with respiration rate when B = 1. Thus, the sampling logic equations are determined to be SHR = LB + L'B' for heart rate, and SRR = LB' + L'B = SHR' for respiration rate. The SRR signal can also be used to control the divider multiplexing circuit as heart rate is processed when the control input signal is a logic 0, while the same is done for respiration rate when the control input signal is a logic 1.

Since the LD130, billed as the first 3 digit monolithic DVM integrated circuit, was introduced in 1976 (Vanderkooi, 1976), analog-to-digital converters have become more sophisticated, making their use as panel meters much easier. The LD130 operates as a self-contained analog-to-BCD converter, with the BCD outputs strobed on a digit basis to directly drive standard CMOS display decoder/driver circuits. More recently the ADD2500 2-1/2 digit A/D (National Semiconductor, 1977a) and the ICL7106 3-1/2 digit A/D (Swift, 1979; Fox, 1980) have been used which include all necessary decoders and drivers. The PCIM176, IC33, is a 3-1/2 digit LCD digital meter module, which uses the dual slope method of A/D conversion (PCI Displays, 1980). In this method, a complete measurement can be divided into three repetitive cycles, autozero, integrate, and read or deintegrate, with the autozero and integrate cycles being of fixed time periods controlled by the master clock, and the read period being of variable length proportional to the unknown input voltage. Features of the PCIM176 include (PCI Displays, 1980):

0.5 inch digit height
200 mV full scale reading
True differential-input and reference
Single 9 V supply

Power dissipation typically less than 20 mW
Guaranteed zero reading for 0 V input
Maximum deviation from best straight line fit ±1 count
Leakage current at input 1 pA
Supply current typically 1 mA

As all design up to this point has been geared to 2 V full-scale, the DPM module is used in the 200 mV digital voltmeter configuration, with a 10:1 input attenuator. Although a 9 V battery is recommended to power the unit, as it is intended that the analog inputs float with respect to the supply voltage, a nonfloating supply can be used at the saving of a battery if care is taken that the analog inputs go no closer than 1.5 V to either supply voltage. To meet this requirement, the manufacturer suggests the use of a split supply or the use of a resistive voltage divider network, though the latter alternative appears to fail in keeping INLO > 1.5 V when the voltage to be measured approaches low levels. An alternative not mentioned, probably due to its increased complexity, would be to use a level shifter to offset the attenuated voltage (Vwiper to ground), with the offsetting voltage used at INLO and the shifted voltage used at INHI. Neither method, split supply or level shifting, is used in this preliminary design.

Alarm system

Audible and visual alarms are frequently utilized, with their important specifications including accuracy, delay in activation, and a method of resetting. Although long time delays reduce false alarms from transient events, with some instruments requiring that the alarm limit be exceeded continuously for 1 to 10 seconds, they delay personnel response to true alarm situations (Fox, 1977). The scheme to be used here, shown in Figure 11, requires that the rate alarm limits be exceeded for four consecutive events before the alarm indicator is activated. Of the two modes generally used to reset alarms, the automatically resetting method is selected, in which the alarm indication is cancelled automatically when the parameter returns to within nonalarm range. A manually reset alarm, on the other hand, would require action by the user to be reset, even if the parameter causing the alarm returns to within nonalarm range, thereby ensuring that no alarm would go unnoticed.

The LM139 series of integrated circuits, IC25 and 29, are four independent precision voltage comparators designed specifically to operate from a single power supply, and like the LM324, have an input common-mode voltage range which includes ground even when the device is operated from a single power supply (National Semiconductor, 1980). The output of the device, the uncommitted collector of a grounded emitter NPN output transistor, can be used as a simple SPDT switch to ground, and allows many collectors to be tied together to provide output ORing. The output pull-

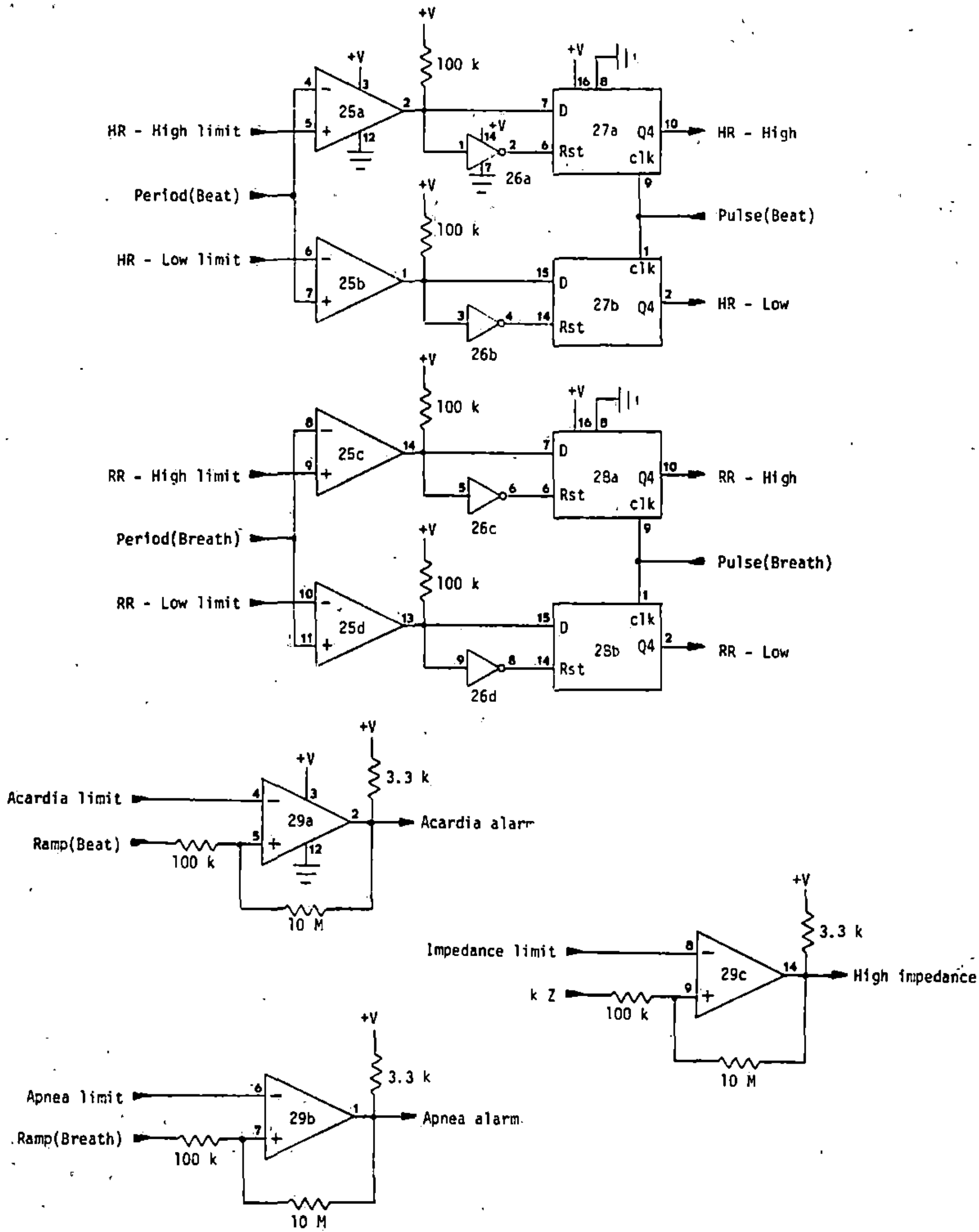


FIGURE 11. Alarm detecting circuit

up resistor should be selected to be large enough so as to avoid excessive power dissipation, and yet be low enough to supply enough drive to switch whatever load circuitry is used on the comparator output (Smathers et al., 1973). Typical characteristics of the LM339 include (National Semiconductor, 1980):

Supply current drain independent of supply voltage	0.8 mA
Output current sink	16.0 mA
Leakage current	0.1 nA

Since comparators are basically uncompensated high-gain operational amplifiers, a small amount of positive feedback or hysteresis is normally added for slowly varying input signals, which force the comparator to stay within its linear region, and thus risk oscillation.

The CD4015 integrated circuit, IC27 and 28, used as the alarm delay element consists of two identical, independent, four-stage serial-input/parallel-output registers (National Semiconductor, 1978). Each of the registers, which consist of four D-type master-slave flip flops, is controlled by independent clock and reset inputs, and has available a single data input and the Q-outputs of each of the four stages. The second monostable multivibrator of the inter-event interval detector serves as the clock input to the shift registers, with the logic level present at the data input, alarm/no alarm, transferred into the first register stage on a positive-going clock transition and shifted over one stage at each subsequent clock. Thus the output, taken from the last stage of the shift register, will go high only if the alarm condition is exceeded for four consecutive events. Resetting of all stages of the register is accomplished automatically by a high level on the reset line, i.e., when the parameter returns to within the alarm limit, and will remain in this state until the comparator limit is again exceeded.

The range of the variable alarm limits and the settings for the fixed alarms are as follows:

• Heart rate - High	100 - 136 beats/min
- Low	43 - 78 beats/min
• Respiration rate - High	24 - 37 breaths/min
- Low	6 - 12 breaths/min
• Acardia	1.5 sec (40 beats/min)
• Apnea	10 sec (6 breaths/min)
• High impedance	855 ohms

Z = Vref/(Ipeak)(G(IA))

As the divider is multiplexed, the corresponding frequency of each inter-event interval is not available; thus, the rate comparators must monitor the period between events. Analysis of the alarm comparators with 1% positive feedback using the methods of Smathers et al. (1973) gives results (shown in Table 2) which show that because of the inverse relationship between period and frequency, $f = 1/T$, positive feedback is not suitable, particularly at the higher frequencies. Reducing the positive feedback of the comparator by reducing the 100 kohm resistor at the positive input terminal increases the problems associated with the loading of the reference voltages, necessitating the use of voltage followers as buffers. Thus, positive feedback is used only on the acardia, apnea, and high impedance alarms, with the potential oscillation of the rate alarms, when the input is close to the limit, resetting the delay shift register.

If a battery-powered system requires a number of LEDs in its display, the power dissipated may be far more than that for all the rest of the circuitry, as each LED requires more than 15 mA of current for a bright display. One way to cut down on current consumption, suggested by Patterson (1974), is to connect all the LEDs in series in a 20 mA current chain, with each LED then controlled by shorting it out with a transistor. Since the transistors can be operated with less than 0.3 mA of base current, normal CMOS logic outputs can provide sufficient current for driving the LEDs, and thus, high current buffers are no longer necessary. In order to avoid excessive emitter-base voltages, the LEDs controlled by the PNP transistors, which allow the LED to turn on for positive logic, are inserted at the top of the chain, while those controlled by the NPN

TABLE 2. Sample thresholds of alarm comparators with 1% positive feedback

alarm parameter	setting/ on/off	volts	events per minute	ohms
Heart rate - High	set	0.25	120.0	
	on	0.247	121.5	
	off	0.336	89.3	
Heart rate - Low	set	0.5	60.0	
	on	0.505	59.4	
	off	0.415	72.3	
Respiration rate - High	set	1.0	30.0	
	on	0.99	30.3	
	off	1.08	27.8	
Respiration rate - Low	set	3.33	9.0	
	on	3.36	8.93	
	off	3.27	9.17	
Acardia	set	0.75	40.0	
	on	0.758	39.6	
	off	0.668	44.9	
Apnea	set	5.0	6.0	
	on	5.05	5.94	
	off	4.96	6.05	
High impedance	set	1.27		855
	on	1.283		864
	off	1.193		803

transistors, which allow the LED to turn on for negative logic, are placed at the bottom. As the number of LEDs desired for the monitor display (14) exceeds what could reasonably be expected to be driven in a single string given a 9 V source, they were separated as shown in Figure 12a.

As LEDs need not be constantly driven to produce the desired visual effect, the scheme shown in Figure 12b reduces the necessary current in half by sending a 20 mA current through the LED strings during alternating periods. The duty cycle (tlow/T) of a standard astable multivibrator circuit can never be equal to 50%, as is desired for an even brightness between the two strings. However, by connecting a diode in parallel with Rb, a duty cycle of 50% or greater can be obtained by allowing the capacitor to charge through Ra and the diode, and discharge through Rb and the discharge terminal (Coughlin and Driscoll, 1977). With C = 0.1 uF and Ra = Rb = 68 kohm, the astable multivibrator, IC24b, has a period of 9.45 ms, which corresponds to a frequency of 106 Hz. It should be noted that the upper resistor value in the figure has been reduced slightly to take into account the voltage drop across the diode when the timing capacitor is being charged, thus balancing the on/off periods of the current source.

The basic concept of a current regulator is to maintain a definite value of current flowing in a variable load impedance, independent of load variations, temperature changes, and supply voltage changes. In the unidirectional current sources used, IC30a and 30b, the load resistance is moved outside the feedback loop of the op amp, placed in the

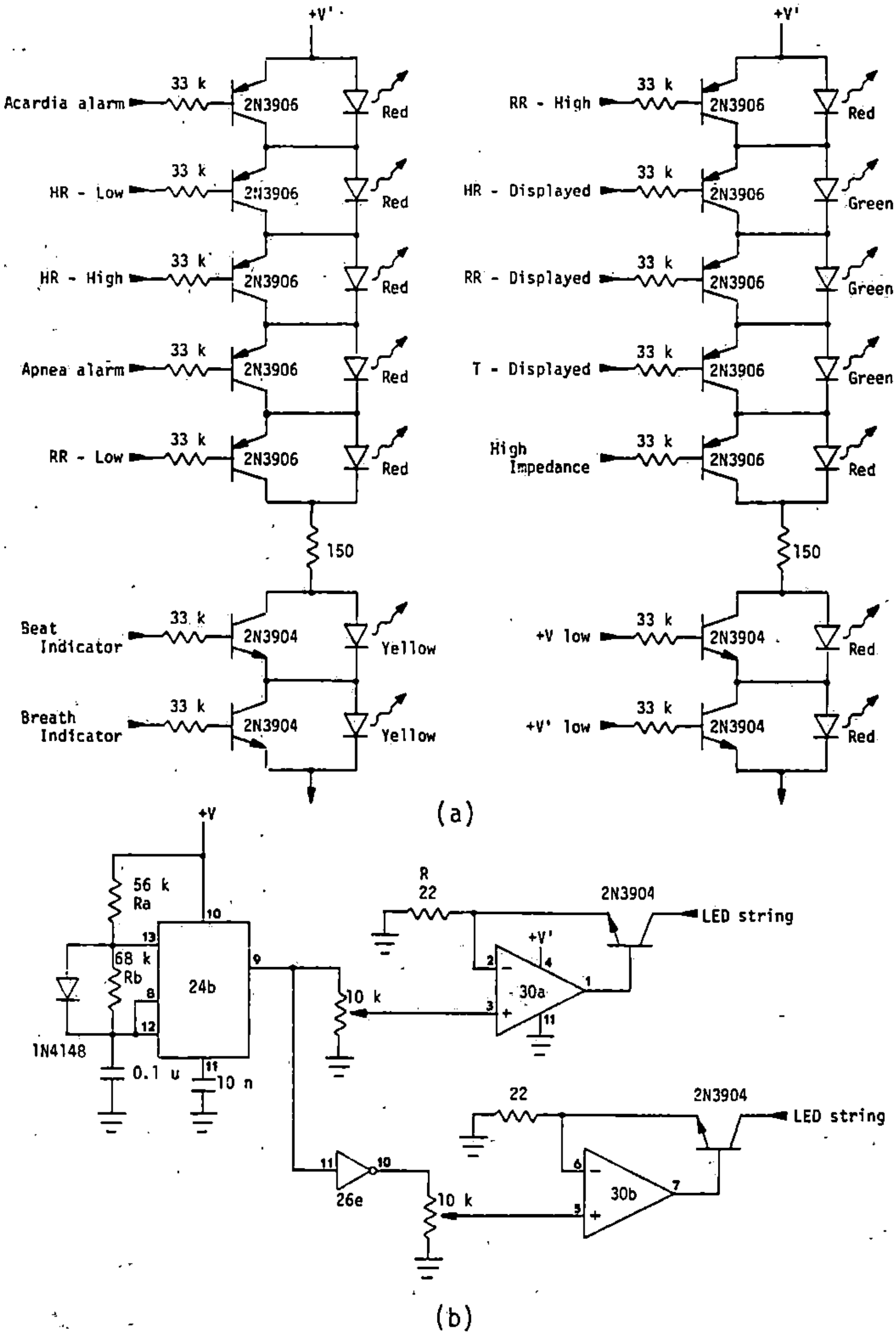


FIGURE 12. Visual alarm indicator using two strings of LEDs (a) and an alternating constant current source (b)

collector circuit of a transistor, and referenced to a more positive supply potential (Jung, 1980). As the current actually regulated by the circuit is the emitter current, which corresponds to Vin/R = Ib + Ic, there is a slight error in the magnitude of the load current. The transistor is selected for the desired operating currents, primarily from the standpoint of power and voltage rating, with a high hFE minimizing the base current error. As designed, the constant current flowing alternately through the LED strings is 22.7 mA (0.5 V/22 ohm).

The audible alarm circuit, shown in Figure 13, is controlled by a multiple input OR gate built of one-fourth of an LM339, IC29d, where a logic 1 at any of the inputs will produce a logic 1 at the output (Smathers et al., 1973). The input diodes can be replaced by resistors whose values are selected such that the positive terminal will be greater than the negative terminal when any of the inputs goes high. However, caution must be taken that the input which goes high is not required to source an excess amount of current and that the positive terminal does indeed exceed the negative terminal reference voltage. For example, if we were to change the diodes to 1 Mohm resistors,

$$V+ = Vin\ (1\ Mohm/7)/(1\ Mohm + (1\ Mohm/7)) = 1.125\ V$$

which is not enough to toggle the output as presently designed, and the negative input terminal reference voltage would need to be reduced to less than 1.125 V.

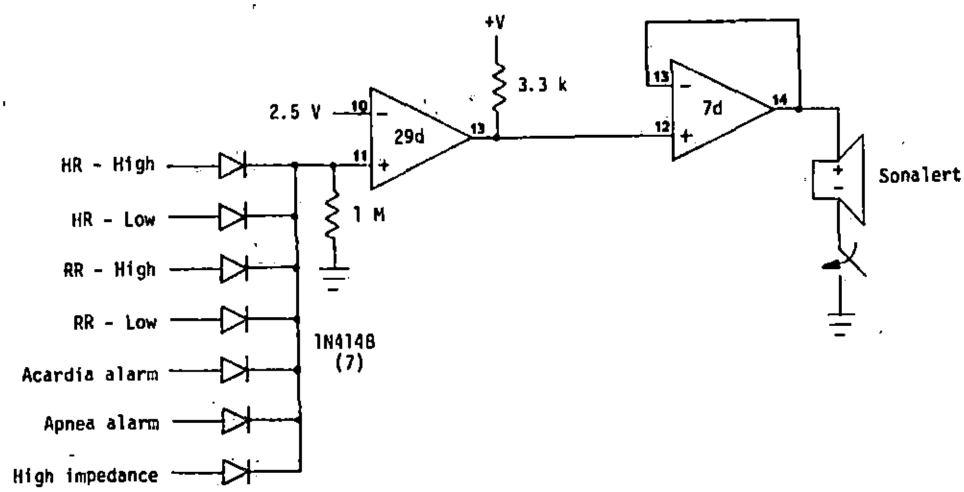


FIGURE 13. Audible alarm indicator circuit

The element used to produce the audible sound denoting an alarm condition is the Sonalert, which has the following characteristics (Jameco, 1982):

Operating voltage	6 to 28 V
Current	3 to 14 mA
Frequency	2900 Hz ± 500
dBA	69 to 80

As the LM339 is capable of sinking a sufficient amount of current, it is possible to save IC7d by replacing the multiple input OR gate with a multiple input NOR gate and placing the Sonalert in series with an appropriate pull-up resistor. The change from an OR gate to a NOR gate simply requires the interchanging of the comparator inputs. An on/off switch is provided to allow the disabling of the Sonalert during calibration or monitor hook-up, though no volume control is included.

Reference voltages

Many of the circuits described previously, i.e., the integrator, the divider, the temperature shifter, and the alarm limits, require constant stable voltage references for proper operation. The circuit shown in Figure 14 utilizes an LM336 precision 2.5 V shunt regulator diode and op amp circuits to produce the required references. This regulator diode operates as a low-temperature-coefficient 2.5 V zener with a 0.2 ohm dynamic impedance, and has a third terminal which allows the reference voltage or temperature coefficient to be trimmed (National Semiconductor, 1980). As shown, the 10 kohm potentiometer permits the adjusting of the reverse breakdown voltage, to take into account initial device tolerance, without affecting the temperature coefficient.

The buffer circuits used are the common positive gain amplifier (IC30c), voltage follower (IC30d), and negative gain amplifier (IC3d), giving outputs of +5 V, +2.5 V, and -3.75 V respectively. The potentiometer value of 1 kohm was selected to be small enough so as to avoid being loaded by the following stage's input, and yet be large enough so as to be within the buffer amplifier's output current ratings.

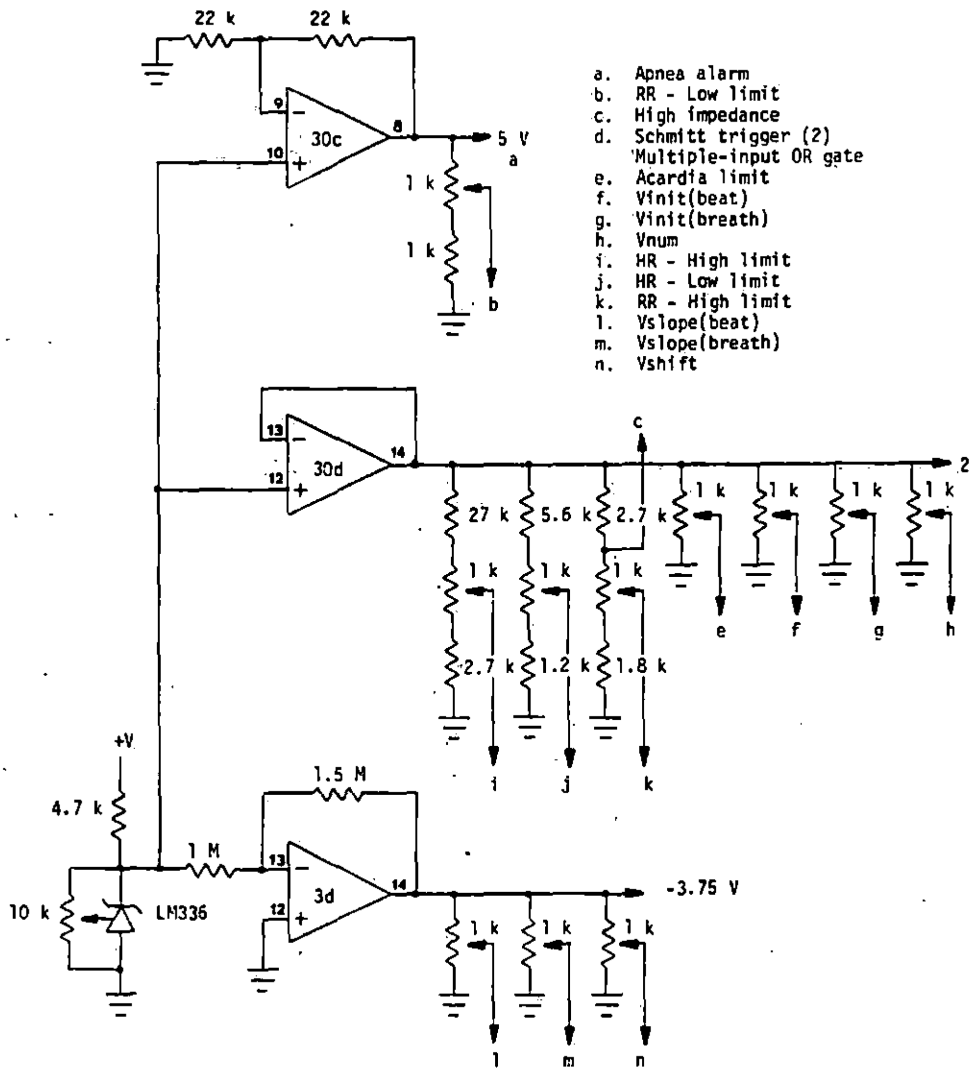


FIGURE 14. Reference voltage circuit

IC30c and 30d source 2.61 mA and 10.86 mA respectively, while IC3d sinks 11.25 mA, all well within the range of the LM324. If higher valued potentiometers are to be used to reduce the buffer amplifier output currents, or destination input impedances reduced, i.e., by introducing hysteresis to the rate alarm comparators, voltage followers may be required to buffer the reference voltage output.

Power supply

The power supply of the monitor, shown in Figure 15, uses four alkaline 9 V transistor batteries, with two used for +V, one for -V, and the last as the floating supply of the DPM. Two +V supplies were deemed necessary as portions of the monitor, i.e., the LED strings, require a proportionately large amount of current with respect to the rest of the circuit.

A battery threshold detector must alert the user when the batteries need to be replaced, but without significantly reducing battery life. In the circuit suggested by Rubinstein (1980), the voltage divider formed by the 1 Mohm resistor and the 2 Mohm potentiometer divides the battery voltage into a value close to the threshold of the CMOS inverter, putting it in its linear operating range. The threshold voltage remains relatively constant with respect to changes in the 9 V battery supply, so that as the battery weakens, the output of the first stage will go high when the input drops below the threshold. To set up the circuit, a voltage equal to the desired cutoff is applied in place of the battery, and the potentiometer adjusted until the output buffer toggles. By properly selecting the resistor and

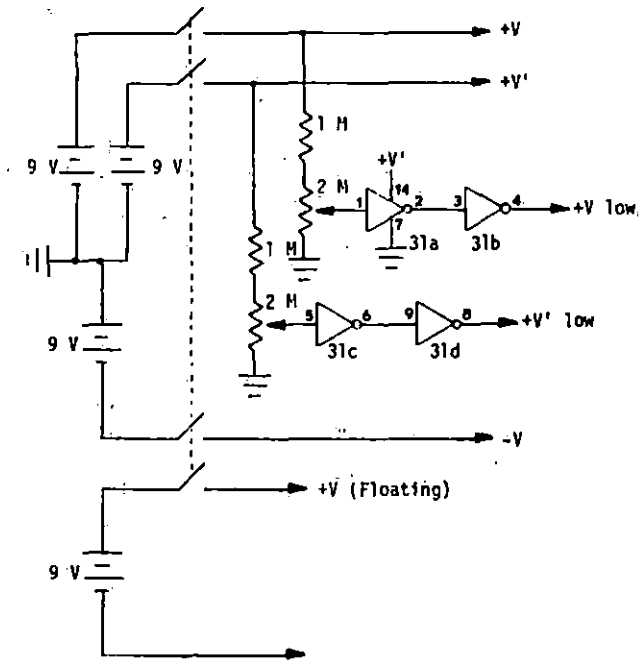


FIGURE 15. Power supply with low-battery indicator

setting the trimmer, the circuit can detect any cutoff voltage that lies between the CMOS threshold and its maximum voltage rating. Only the two +V supplies are monitored as they are the ones expected to have the highest current drains. As presently set, the low-battery LEDs will turn on when these battery voltages go below 8.5 V, being slightly dimmer initially compared to the other LEDs, due to the oscillatory output.

Arterial Blood Pressure Parameters/Pulse Rate Monitor

The circuit diagram of the front-end of the Arterial Blood Pressure Parameters/Pulse Rate monitor is shown in Figures 16 through 18, with the integrated circuits used listed in Table 3. Features of this monitor include:

- Direct systolic pressure
- Choice of direct diastolic or mean pressure
- Beat-to-beat pulse rate
- 3-1/2 digit LCD display
- Two display modes, cycle and select, with LED indicating parameter being displayed
- Audio and visual alarm indication, with audio disable
- Adjustable high/low rate and pressure alarms
- Acardia alarm
- Pulse indicator
- Battery operation with low-battery indicator

Transduction and calibration of the arterial pressure waveform

The circuit used to convert the arterial pressure waveform, from a mechanical variable to an electrical one, is shown in Figure 16. Included in the LX0503/LX0603 series of monolithic piezoresistive integrated circuit pressure transducers is the LX0603GB, IC8, a gage transducer which is

TABLE 3. IC labels used in Figures 16 through 18 and their corresponding integrated circuits

LABEL	IC	LABEL	IC
1	LM324	5	LM308
2	LM324	6	LM308
3	LM308	7	LM308
4	LM308	8	LX0603GB

suitable for use with nonacidic working fluids including water (National Semiconductor, 1979a). These pressure transducers include only the basic monolithic pressure IC chip used in National's hybrid transducers, and do not have the scaling and buffering circuitry, single-ended high-level output, or offset temperature compensation of the hybrid devices. Although effective application of the monolithic device will require external circuits to perform the above functions, these devices have the advantages of greatly reduced unit cost and increased design flexibility. The device does have a built-in span temperature compensation circuit which is adequate for most users, requiring only that VE be regulated, repeatable with temperature. Guaranteed characteristics for 0 to 30 psi include (National Semiconductor, 1979a):

Offset calibration ±100 mV at 0 psi

Offset stability	±0.5 psi
Sensitivity	0.038 to 0.154 mV/mm Hg
Linearity	±0.4 psi
Span stability	±0.2 psi
Bias current	5 mA
Bridge resistance	1800 ohms
Natural frequency	50 kHz

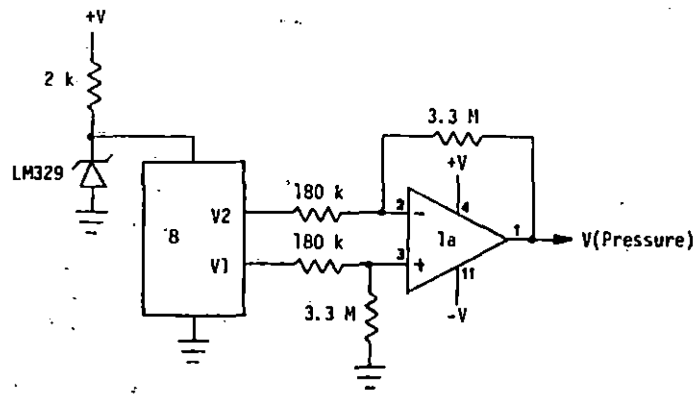


FIGURE 16. Arterial blood pressure waveform transduction circuit

The heart of the transducer is a monolithic silicon chip with a cavity etched out to form a diaphragm, the top side of which contains the pressure sensing circuitry. The LX0603 series of transducers have two pressure inlets which differ in susceptibility to moisture and other fluids. The ambient pressure port allows the working fluid to make contact with the circuit side of the diaphragm which is covered with a thin compliant layer of parylene. While parylene protects the circuitry in a humid environment, it does not protect against water and other aqueous and ionic fluids, and appropriate precautions must be taken. On the other hand, the working fluid port makes contact with the cavity side of the diaphragm which is insensitive to water and ionic fluids.

The output of the transducer is taken directly from the diffused piezoresistive Wheatstone bridge; thus, the interface is a critical step of signal conditioning. If designed and fabricated properly, with little or no bridge loading by external resistors, the interface buffer circuit will provide high gain and minimum distortion of the transducer characteristics and temperature coefficients. The most effective bridge buffering circuits use difference amplifiers having high impedance, well-matched, carefully installed resistors, with improved performance obtained using high quality instrumentation amplifiers or dual voltage followers. The gain of the difference amplifier shown, IC1a, is 18.3.

Having obtained a voltage proportional to the arterial pressure waveform, the circuit shown in Figure 17 is used to filter out all high-frequency noise and calibrate the signal. In a low-pass filter, all input signals whose frequency is within the pass-band of DC and the breakpoint are transmitted, while those with frequencies greater than the breakpoint are attenuated. A common filter approximation uses the Butterworth polynomials, and has a transfer function of the form (Millman and Halkias, 1972):

$$Av(s)/Avo = 1/Bn(s)$$
$$= 1/((s/wo)^2 + 2k(s/wo) + 1),$$

where k = damping factor

wo = -3 dB frequency

= 1/RC

Avo = stable midband gain

= 3 - 2k = 1 + (Rf/R).

The typical second-order filter to be used has a -40 dB/decade attenuation and the following transfer equation (Millman and Halkias, 1972):

$$Av(s)/Avo = 1/((s/wo)^2 + 1.414(s/wo) + 1).$$

Thus, k = 0.707 and Avo = 1.586. In order to pass the 10th harmonic of a 200 beat/minute signal, the pass-band must be 33.3 Hz, which corresponds to an RC time-constant of 4.78 ms. As designed, with filter resistors and capacitors of 47 kohm and 0.1 uF respectively, the low-pass filter, IC1b, has a -3 dB point at 33.9 Hz, and a gain of 1.56.

The principal problem encountered in the calibration and scaling of a transducer is the interaction of offset (common-mode) and span (normal-mode) and their respective temperature coefficients. For easy calibration and scaling, the simplest technique for effective reduction requires the use of two amplifier stages. The first, a noninverting

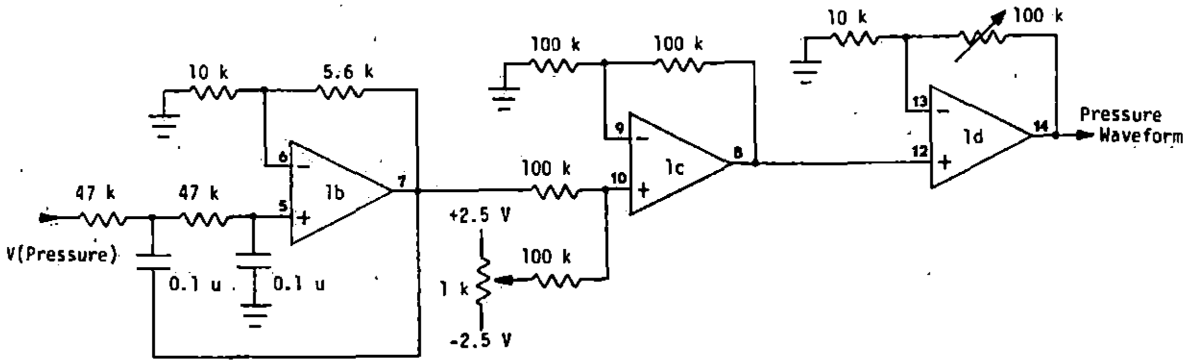


FIGURE 17. Pressure waveform filtering, balancing, and scaling circuit

summer, IC1c, serves to balance the transducer offset voltage, causing Vout to be zero at an ambient input pressure. The second, a noninverting amplifier, IC1d, serves to scale the transducer output, and has a gain which ranges from 1 to 11. With the transducer sensitivity ranging between 0.038 and 0.154 mV/mm Hg, the gain required to obtain the desired sensitivity of 10 mV/mm Hg would be 263.2 and 64.9 respectively. As the difference amplifier has a gain of 18.3, and the low-pass filter a gain of 1.56, the gain required of the noninverting amplifier is thus the range 9.22 to 2.27.

<u>Determination of the arterial pressure parameters</u>

Having properly calibrated and scaled the voltage waveform corresponding to the arterial pressure, the parameters of mean, systolic, and diastolic can be determined using the circuit shown in Figure 18. The time-weighted average, or mean pressure, is determined by using a second-order Butterworth low-pass filter, IC2a, with a severely restricted pass-band, 0.072 Hz as shown, corresponding to a time-constant of 2.2 seconds. This value is selected to be a decade smaller than the minimum heart rate expected, i.e., 45 beats/minute or 0.75 Hz, so that at this minimum rate, only 1% (-40 dB) of the pressure waveform variation will be passed to the output. As the Butterworth filter has an inherent gain associated with it, a potentiometer with a voltage follower is used to properly scale its output.

As the determination of systolic pressure is a process similar to that of determining the change in the carrier wave caused by variations in thoracic impedance, a similar peak detector circuit is used. As the frequency of the pressure waveform is a great deal less than that of the carrier wave, the clamping diode used to speed-up the LM308 response is no longer necessary, however, greater care must be taken in selecting the storage capacitance value. When extremely long holding times are required of a peak

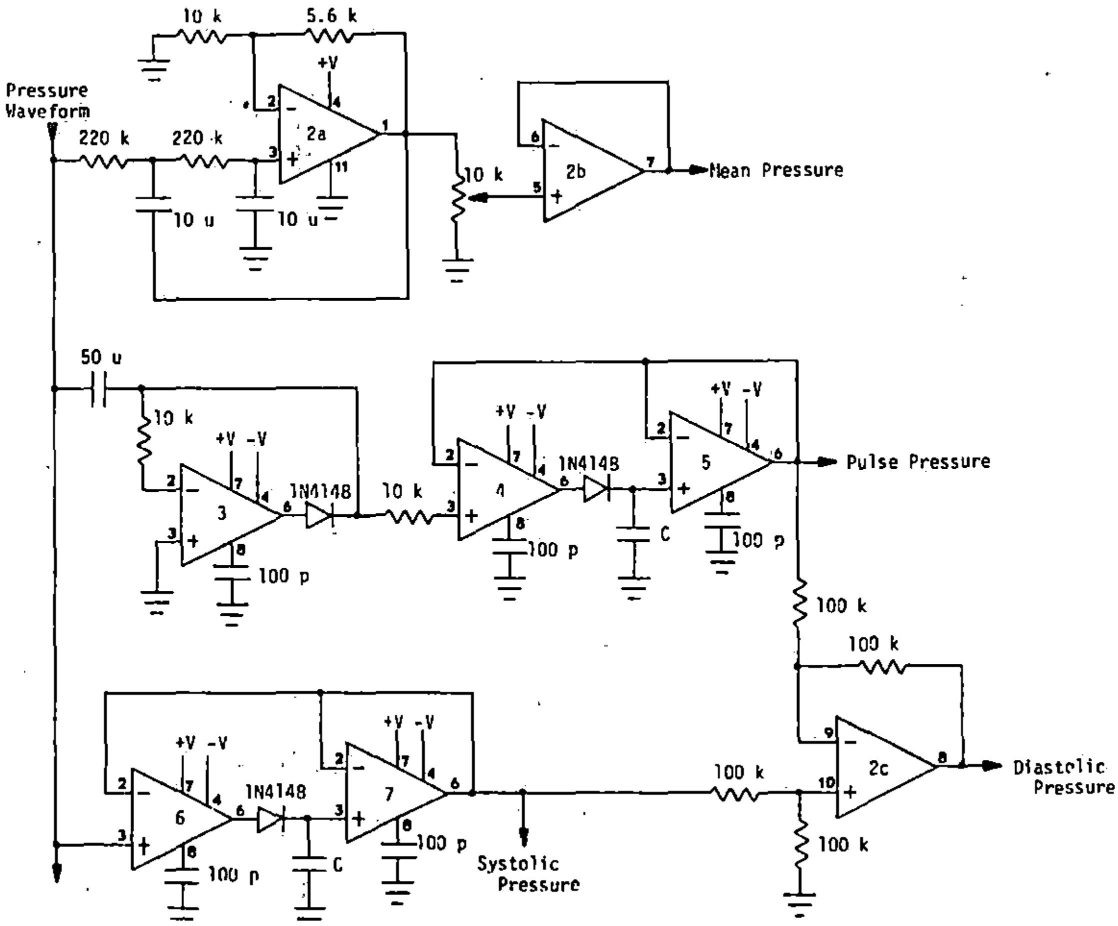


FIGURE 18. Circuit to determine systolic, diastolic and mean pressures

detector, the discharge current of the capacitor becomes important, and is thus buffered by a low-input-current voltage follower (Jung, 1980). In practice, the diode leakage current will play a significant role and must be taken into account when selecting the capacitance value.

The equation controlling the change in capacitance voltage, $dV/dt = I/C$, deems that a large value of C will help maintain the capacitance voltage, but a tradeoff exists as a large value of C will slow down the response of the peak detector when the input peak value changes rapidly. Although silicon diodes have a 0.6 V threshold which must be overcome before appreciable conduction occurs, placing the diode in the feedback loop of an op amp, causes the threshold voltage to be divided by the open-loop gain of the amplifier and virtually eliminated, causing very little error in the peak value detected (Millman and Halkias, 1972).

The determination of diastolic pressure is slightly more complicated, as it is obtained by subtracting the pulse pressure from the systolic value. To obtain the pulse pressure value, the pressure waveform is capacitively coupled to a precision clamp, IC3, forcing the most negative point of the waveform to ground. Its peak value, which corresponds to the desired pulse pressure, is then determined by using a positive peak detector of the same type used to determine systolic pressure. In order to clamp the input waveform's negative peak at Vref, the precision clamp circuit begins by allowing the coupling capacitor to charge to a voltage that equals the mean of the input (Jung, 1980). When the negative input terminal swings below Vref, the op amp drives the diode on, forming a low impedance path to charge the coupling capacitor in the opposite direction, with the net effect being to discharge it. After the peak has passed and the diode is again turned off, the charge on the coupling capacitor will be retained if a large enough capacitance value is used, thus giving a fast attack, slow decay response to input level changes. As with the precision diode, the high gain of the operational amplifier allows this circuit to be responsive into the uV region, causing very little change in the amplitude of the input waveform. With the detection of the positive peak of the clamp output, the diastolic value can be determined using a difference amplifier, IC2c, with a gain of one, the systolic value being the positive input and the pulse value the negative.

The remainder of this monitor uses circuits essentially the same as that described above for the Heart Rate/Respiration Rate/Temperature monitor, though a few modifications are necessary, including:

- Pulse rate can be determined by using the pressure waveform as the input to the circuit used in determining heart rate above, beginning with the noninverting amplifier/Schmitt trigger circuit, though in this instance the analog divider need not be muliplexed.

- As the systolic and diastolic pressures should be input to sample-and-hold circuits to stabilize the display, the sample-and-hold logic must be increased in order to be able to sample the three parameters at the appropriate times.
- The rate alarms can now use positive feedback, as the parameter to be monitored would be the frequency and not the period.
- High- and low-pressure limit alarms of the same configuration as that used for rate should be included for systolic and diastolic/mean.
- A selector switch for diastolic/mean should be included such that the desired parameter is input to the DPM and the alarm circuit.

FABRICATION, CALIBRATION, AND TESTING

Because much of each monitor is composed of the same circuitry, it was decided that the complete Heart Rate/Respiration Rate/Temperature monitor would be assembled and packaged, while only the waveform separator of the Pressure Parameter/Pulse Rate monitor would be breadboarded and tested.

Heart Rate/Respiration Rate/Temperature Monitor

In order to facilitate the modifications that would be required in the course of the monitor development, the circuits were fabricated by wire-wrapping. All components with leads that could be easily inserted in wire-wrap sockets were so assembled, while those that could not, because of lead size or configuration, were inserted directly into the perforated board, wire-wrapped, and soldered. The board was sized such that when housed in the molded plastic case, space was available for battery placement and front panel leads, with an attempt made to arrange the sockets such that the lead lengths - from socket to socket and board to panel - were kept to a minimum. On the aluminum front panel of the monitor were placed all LED indicators, the on/off switches for power and the audible alarm, a push button switch for cycle/select, the two sensitivity adjustments, the four sliding alarm limit adjustments, a three-connector jack for the electrode leads, a two-connector jack for the temperature probe, and the digital panel meter. The Sonalert was placed on the aluminum back panel of the case.

With all circuits wire-wrapped and leads tested to ensure proper connection between terminals, the monitor was adjusted and calibrated. Among the first things to be adjusted were:

- Sine wave amplitude (Figure 1) - Vout = 8 V peak-to-peak
- Instrumentation amplifier output offset (Figure 2)
- LED current source (Figure 12) - Vin = 0.5 V
- LM336 trim (Figure 14) - Vref = 2.5 V
- Low battery indicator (Figure 15) - Vtoggle = 8.5 V

Disconnecting the waveform separation and impedance change demodulation circuit, Figure 3, from the amplification and triggering circuit, Figure 4, allowed for easy calibration of the inter-event interval detectors (Figures 5 and 6), the analog divider (Figure 8), and the digital panel meter (Figure 10). The procedure used was as follows:

1. Input a known constant period waveform to the heart beat amplification and triggering circuit.
2. Adjust the first monostable multivibrator to obtain a period of 0.15 seconds.

3. Monitor the output of the NOR gate and adjust the second monostable multivibrator to get a period of 0.3 seconds
4. Monitor the output of the integrator and adjust Vinit so that the reset voltage is 0.15 V.
5. Still monitoring the output of the integrator, adjust Vslope so that the slope during integration is 0.5 V/sec and the voltage held after integration is (0.5 V/sec)(Tin).
6. With the integrator output known, determine the analog divider output required, 0.3/Vint, and adjust Vnum to obtain the proper results.
7. Finally, adjust the digital panel meter attenuator so that the panel reading is equal to the input frequency.

Steps 1 through 5 were then repeated for the respiration rate circuit. Note that the heart rate monitor was calibrated first, as the short interval between events may cause the offset voltages of the divider to result in significant errors. Both monitors were calibrated with an input frequency of 100 events per minute.

To calibrate the temperature monitor, the following procedure was used:

1. Immerse the probe into a water bath of known temperature.

2. Monitor the output of the voltage follower and adjust the potentiometer so that Vout = (10 mV/°C)(Tbath) + 2.7316 V.
3. Monitor the output of the summing amplifier and adjust Vshift so that Vout = (10 mV/°C)(Tbath), and the panel reads the correct temperature of the water bath.

The temperature monitor was calibrated in a 37°C water bath.

Testing of the rate meters was accomplished by varying the period of the input waveform, with the results obtained for the heart rate monitor shown in Table 4, and that for the respiration rate monitor shown in Table 5. It should be noted that the absolute errors are shown, rather than the more commonly used relative errors, for reasons that will become evident below. Having also recorded the output voltage of the integrator sample-and-hold circuit, we are able to localize the major sources of error, as follows:

- Knowing the period of the input waveform allows the anticipated sample-and-hold output voltage to be calculated, (0.5 V/s)(Tin), and the errors of the inter-event interval detector determined.
- Knowing the sample-and-hold voltage of the monitor allows the anticipated digital panel meter value to be calculated, 0.3/Vsh, and the errors of the analog divider determined.

TABLE 4. Heart rate monitor panel readings with varying input periods

Input period (sec)	Calculated frequency (per min)	Panel reading (per min)	Absolute error (per min)
0.600	100.0	100.0	0.0
1.508	39.8	36.4	-3.4
1.208	49.7	46.1	-3.6
0.990	60.6	57.0	-3.6
0.838	71.6	68.4	-3.2
0.734	81.7	79.3	-2.4
0.676	88.8	87.3	-1.5
0.542	110.7	112.3	1.6
0.501	119.8	122.5	2.7
0.462	129.9	134.7	4.8
0.428	140.2	147.7	7.5
0.402	149.3	159.3	10.0
0.376	159.6	172.8	13.2
0.353	170.0	186.8	16.8
0.336	178.6	199.3	20.7

Results of the above procedure are shown in Tables 6 and 7 for the heart rate and respiration rate monitors, respectively. In examining the errors of the heart rate monitor, we can conclude that although the reset of the integrator was too big and the slope too small, most of the error obtained was due to errors in the analog divider conversion, i.e., compare the frequency errors listed with the value obtained by multiplying the divider error by 100. In the case of the respiration rate monitor, the errors of

TABLE 5. Respiration rate monitor panel readings with varying input periods

Input period (sec)	Calculated frequency (per min)	Panel reading (per min)	Absolute error (per min)
0.600	100.0	100.0	0.0
9.92	6.0	5.8	-0.2
5.98	10.0	8.7	-1.3
4.31	13.9	12.0	-1.9
3.39	17.7	15.4	-2.3
2.75	21.8	18.9	-2.9
2.32	25.9	22.6	-3.3
2.020	29.7	26.1	-3.6
1.784	33.6	29.8	-3.8
1.596	37.6	33.5	-4.1
1.444	41.6	37.3	-4.3
1.320	45.5	41.2	-4.3
1.204	49.8	45.3	-4.5
1.124	53.4	49.0	-4.4
1.048	57.3	52.9	-4.4
0.976	61.5	57.2	-4.3

the analog divider again dominate, with the relative errors of the period detector generally less than 1% except at 6 breaths per minute where the zener diode begins to take effect. Thus, in future implementations of this rate monitoring scheme, the analog divider must be trimmed for both gain and offset due to the low voltage levels in use.

The temperature monitor was tested by varying the temperature of the water bath, with the results obtained

TABLE 6. Calculations for the determination of heart rate monitor error sources

Input period (sec)	V(Per) calc (mV)	Sample-and-hold (mV)	Abs error (mV)	V(Div) calc (V)	Panel reading (V)	Abs error (V)
0.600	300.0	300.0	0.0	1.000	1.000	0.000
1.508	754.0	742.0	-12.0	0.404	0.364	-0.040
1.208	604.0	591.0	-13.0	0.508	0.461	-0.047
0.990	495.0	492.0	-3.0	0.610	0.570	-0.040
0.838	419.0	414.0	-5.0	0.725	0.684	-0.041
0.734	367.0	365.5	-1.5	0.821	0.763	-0.058
0.676	338.0	336.5	-1.5	0.892	0.873	-0.019
0.542	271.0	269.0	2.0	1.115	1.123	0.008
0.501	250.5	252.0	1.5	1.190	1.225	0.035
0.462	231.0	232.0	1.0	1.293	1.347	0.054
0.428	214.0	214.0	0.0	1.402	1.477	0.075
0.402	201.0	206.0	5.0	1.456	1.593	0.137
0.376	188.0	191.5	3.5	1.567	1.728	0.161
0.353	176.5	180.5	4.0	1.662	1.868	0.206
0.336	168.0	173.0	5.0	1.734	1.993	0.256

shown in Table 8. There appears to be a slight error in the gain factor, which can be attributed to the fact that the digital voltmeter used at the voltage follower output could read only three significant digits, i.e., 3.10 V instead of the desired 3.1016 V. The accuracy of the mercury thermometer used as the reference also probably played a role.

TABLE 7. Calculations for the determination of respiration rate monitor error sources

Input period (sec)	V(Per) calc (V)	Sample-and-hold (V)	Abs error (V)	V(Div) calc (V)	Panel reading (V)	Abs error (V)
0.600	0.30	0.30	0.00	1.000	1.000	0.000
9.92	4.96	4.49	-0.47	0.067	0.058	-0.009
5.98	2.99	3.01	0.02	0.100	0.087	-0.013
4.31	2.155	2.175	0.020	0.138	0.120	-0.018
3.39	1.695	1.705	0.010	0.176	0.154	-0.022
2.75	1.375	1.380	0.005	0.217	0.189	-0.028
2.32	1.160	1.170	0.010	0.256	0.226	-0.030
2.020	1.010	1.020	0.010	0.294	0.261	-0.033
1.784	0.892	0.904	0.012	0.332	0.298	-0.034
1.596	0.798	0.798	0.000	0.376	0.335	-0.041
1.444	0.722	0.716	-0.006	0.419	0.373	-0.046
1.320	0.660	0.655	-0.005	0.458	0.412	-0.046
1.204	0.602	0.603	0.001	0.498	0.453	-0.045
1.124	0.562	0.561	-0.001	0.535	0.490	-0.045
1.048	0.524	0.525	0.001	0.571	0.529	-0.042
0.976	0.488	0.488	0.000	0.615	0.572	-0.043

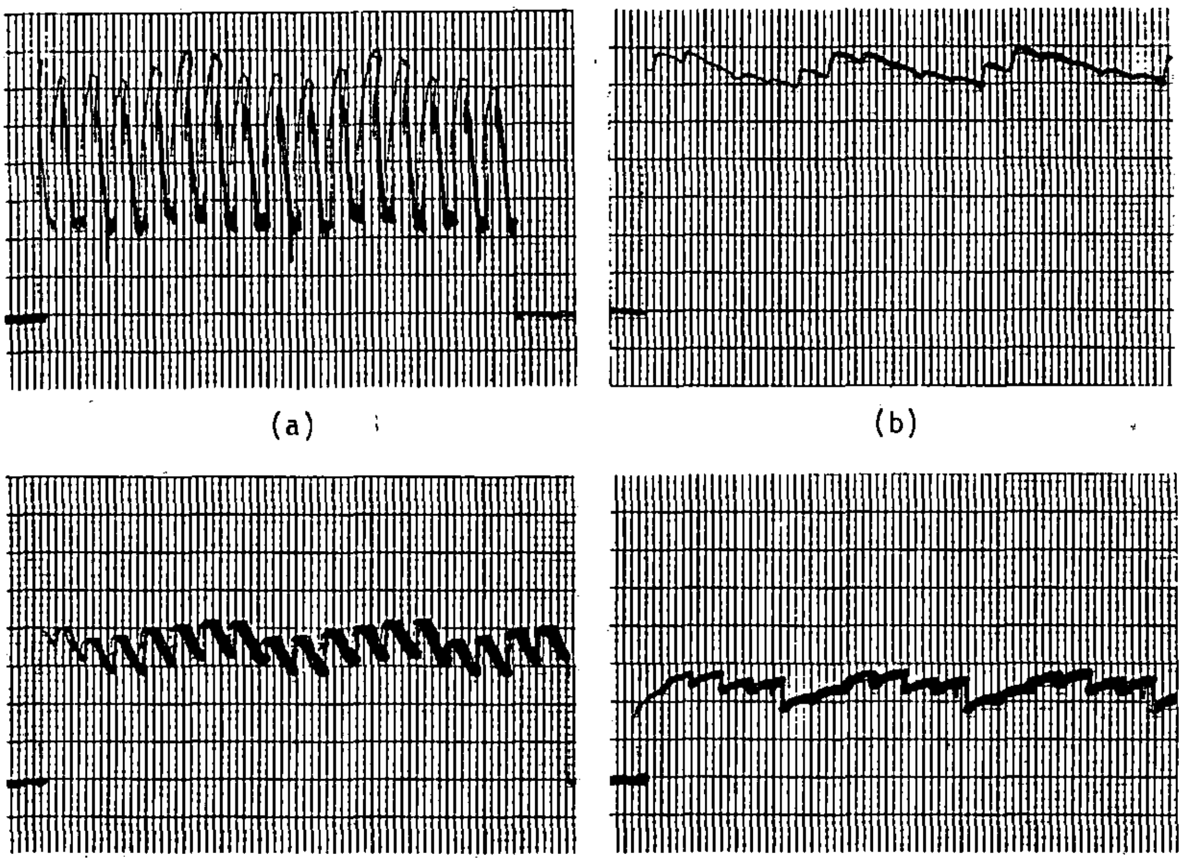
Pressure Parameters/Pulse Rate

The circuits which make up the arterial pressure waveform separator portion of the Pressure Parameters/Pulse Rate monitor, Figure 18, were breadboarded on a superstrip and tested using a pressure waveform simulation device. The only calibration necessary in this portion of the monitor is that of the mean pressure filter attenuator, which can be accomplished by inputting a known DC voltage to the filter,

TABLE 8. Temperature monitor readings with varying water bath temperatures

T(Bath) (°C)	Follower output (V)	Panel reading (°C)	Absolute error (°C)
37.0	3.10	37.0	0.0
35.0	3.08	34.7	-0.3
36.0	3.09	35.9	-0.1
38.0	3.11	38.1	0.1
39.0	3.12	39.1	0.1
40.0	3.13	40.3	0.3
41.0	3.14	41.2	0.2
41.7	3.15	41.9	0.2

and adjusting the potentiometer until the voltage follower output becomes the same magnitude. A sample recording of the input pressure waveform, and the systolic, pulse, and diastolic pressure waveforms obtained is shown in Figure 19. To illustrate the role played by the peak detector hold capacitor, the values used in the systolic and pulse pressure peak detectors were 0.475 and 0.333 uF, respectively.



(a) (b) (c) (d)

FIGURE 19. Pressure waveform separator: input (a), systolic with C = 0.475 uF (b), pulse with C = 0.333 uF (c), and diastolic (d)

SUGGESTIONS FOR FUTURE DEVELOPMENT

Several design alternatives and methods to improve the monitors have been mentioned above in the discussions on circuit development and testing. The following are a few ideas which may be worth incorporating into future implementations.

One of the shortcomings of the present monitor design is that the rate displayed corresponds to the previous inter-event interval, which is held even when the triggering stops. This problem results from the fact that the rate output is updated only after sensing a triggering event, and hence, if no triggering event is detected, no updating can occur. Providing apnea and acardia alarms somewhat circumvents this inherent characteristic, but a better solution would be to make the integrator's sample-and-hold circuit hold its output only when the input voltage is lower than that stored. This scheme would thus make the value displayed always current, by causing the sample-and-hold output to follow the input voltage after a time equal to the last measured period, until the next triggering event occured.

As stated in the design objectives, all circuits utilize ready-made integrated circuit subsections wherever possible. However, the current requirements of a few ICs, such as the sample-and-hold and the analog divider, make their use in the monitor rather expensive. Some of the alternatives possible for prolonging monitor operation using a given set of batteries include:

- The use of rechargeable batteries, such as the nickel-cadmium or gelled electrolyte lead acid batteries, in place of the alkaline batteries which are used once and discarded. These rechargeable batteries have a higher initial cost than the alkaline batteries but can generally be used through hundreds of charge/discharge cycles, with the lead acid battery having a higher energy storage capacity than the alkaline battery, which in turn has a higher storage capacity than the nickel-cadmium battery. It should be noted, however, that to maintain electrical isolation of the patient, charging of the batteries should not take place while the monitor is in use.
- The use of discrete and low-powered integrated circuits, with appropriate device characteristics, to build replacements for the high powered ready-made subsections.
- The use of CMOS [the ICL76XX family (Intersil) and the CA3130/CA3140 (RCA)], programmable [the LM4250 (National)], or micro-powered [the LM10 (National)] operational amplifiers where the device

characteristics meet the circuit requirements.

- The use of large valued potentiometers in the voltage reference circuit buffered by low-power op amp voltage followers.

To protect the pressure transducer against rough handling and to allow the use of a three-way stopcock, it is advisable to mount the transducer in a plastic housing having cable and pressure inlets. The stopcock can then be used to facilitate such things as flushing the line, introducing drugs, zero-referencing to atmosphere, and withdrawing blood samples. The clotting of blood in the catheter tip can be prevented by introducing a continuous flow of a saline solution (usually heparinized) through a micro-pore in the transducer-catheter coupling, thus allowing the system to be perfused automatically without degrading the dynamic signals. It may also be desirable to fill the transducer and pressure inlet with neutral oil to avoid problems of corrosion, sterilization, and air bubbles, though care must be taken that the diaphragm used to separate the oil from the saline solution does not alter the system characteristics appreciably.

Finally, the monitoring of physiological parameters is typically accomplished by a number of separate transducers and instruments surrounding and attached to the patient, cluttering the area and inconveniencing personnel. A possible solution to this problem is the use of an esophageal multiprobe to monitor temperature, heart and lung sounds, and the ECG waveform (Demer et al., 1978). In the design suggested, stainless steel electrodes for monitoring the ECG and a thermistor for monitoring temperature are supported by a small diameter plastic listening tube with integral conductive plastic wiring. Newbower and Cooper (1979), however, warn of the potential hazards of this system which include myocardial stimulation from leakage currents through the electrodes, mechanical injury to the esophagus from hard metal electrodes, and burns due to the passage of currents from electrosurgical devices.

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