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Title: AN ANALYSIS SYSTEM FOR THE REMOTE DETERMINATION OF OXYGEN WITH DATA TRANSMISSION VIA PULSE DURATION TELEMETRY

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Harry Freund

An instrument was developed which performs analysis of the oxygen content of air by means of a commercial polarographic oxygen electrode. The oxygen analyzer is part of a portable instrument which converts the current from the oxygen electrode into a pulse whose width is proportional to the oxygen concentration. This pulse is suitable for transmission by FM telemetry. The instrument is capable of handling up to six additional channels of pulse-width information. The seven channels, plus a channel of synchronizing information, are transmitted sequentially over a radio or wired link to demultiplexing circuitry, which separates the pulses into different channels. The pulse width is then converted into a voltage analog suitable for recording.

The system shows no detectable drift over long periods of operation other than that due to the temperature coefficient of the

oxygen electrode. Linearity of the instrument is excellent except in the pulse-width to analog voltage converter, which exhibits a slightly curved response. As a result, the plot of oxygen concentration at the electrode versus voltage output to the recorder is not quite a straight line. For changes of less than 5 or 10% oxygen, the nonlinearity is not detectable on a recorder.

An Analysis System for the Remote Determination of Oxygen with Data Transmission via Pulse Duration Telemetry

by

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INTRODUCTION

Statement of Problem

The human body requires energy to function. This energy is derived from chemical processes taking place within the body. The reactions which actually produce energy when and where it is needed are complex and not thoroughly understood, but the essential process is the oxidation of organic molecules. A description of the chemistry can be found in any physiology text (3, 15).

The preferred process is the oxidation of carbohydrates such as glucose:

$$C_mH_{2n}O_n + mO_2 \longrightarrow mCO_2 + nH_2O$$

In this reaction, the ratio of carbon dioxide produced to oxygen consumed is 1:1. This ratio is called the "respiratory quotient" or "RQ."

If carbohydrate material is not available for oxidation, then fats are used instead. The complete oxidation of fats requires more oxygen relative to the carbon dioxide produced. The RQ falls to about 0.7. For the oxidation of protein to carbon dioxide, water, and urea, the RQ is about 0.8. The respiratory quotient then, provides an indication of the kind of material that is being used by the body for energy production.

The standard method for measuring the respiratory quotient of an individual is to collect the air he exhales in a container, and, after a period of time, measure the volume, percent carbon dioxide, and percent oxygen. There are several problems associated with such a procedure.

The most obvious limitation results from the size and weight of the gas collection equipment. The individual under test breathes into a large tube which is connected to the gas collection apparatus. Consequently, the individual cannot be far removed from the apparatus, and the apparatus cannot move to the individual. Testing must be done in a laboratory. For individuals at rest, this limitation presents no problem, but for measurements made during exercise, artificial kinds of work must be devised. At Oregon State University, bicycle ergometers are used in the laboratory, but measurements made under more traditional kinds of work conditions, such as running, are impossible.

Another difficulty with the gas collection method is that it does not really measure the right parameters. Gas exchange occurs only in the depths of the lungs, and it is in this alveolar air that the CO_2/O_2 ratio is significant. But the air collected from exhalation includes large amounts of air that never reached the lungs; air that was held in the mouth, esophagus, and bronchii. Meaningful RQ measurements must be made on the air from the depths of the lungs only; this is the last air exhaled in each breath.

What is required is a small portable system which will analyze the exhaled air for carbon dioxide and oxygen, and then transfer the analytical information to another location where it can be recorded on a strip chart recorder. The system should be relatively inexpensive and capable of operating under conditions of physical stress. Since at least two different kinds of information must be handled, the possibility of devising an expandable system should be considered. It might not be much more difficult to handle four channels of information than to handle two channels. Other kinds of information which might be desired include blood pressure, pulse rate, breath rate, temperature, and volume of air respired.

Approach to the System

The analytical problem has two aspects which must be considered. First, methods must be found to conduct the actual analysis of the exhaled air. Second, a system must be devised which will move this analytical information to a central point where it can be recorded. Each aspect of the problem must be considered in light of the other. The data-handling equipment would not be designed to respond to variable frequency sine waves, because the analysis would not be expected to produce such a function. Similarly, fast scan polarography would not be a suitable analytical technique because of the complexity of the associated instrumentation and the difficulty of electronically reducing the data. So, although the two areas of concern will be treated separately, it is important to remember that a unified system with compatible members must be the end result.

If one were faced with instrumental analysis of a mixture of gases in the laboratory, the first choice of method might be mass spectroscopy, but such a technique can only be incorporated into a lightweight instrument at great cost. Another good laboratory method is gas chromatography, but this violates two of the restrictions on the system. It requires fairly bulky equipment and the data is difficult to handle electronically. Other methods, such as paramagnetic resonance for oxygen and infrared absorption for carbon dioxide, can be ruled out because they require delicate equipment. The methods used must be capable of operation under hostile conditions. A method of analysis for oxygen which seems to meet all the requirements is the polarographic oxygen electrode. It is not exceedingly expensive (about $100), it produces an electrical output which is easy to handle, and it is quite rugged.

Carbon dioxide presents a more difficult problem. An electrochemical cell has been used for dissolved CO_2 analysis in body fluids by Stow (21), and a similar technique might be applied to CO_2

in gases. The method relies on sensing changes in pH due to the formation of a bicarbonate-carbonic acid system from CO_2 diffusing across a rubber membrane into an electrolyte. The electrolyte is in a thin film between the rubber membrane and the surface of a glass electrode. The chief difficulty would appear to be the long time required to equilibrate the carbonic acid level with the CO_2 partial pressure on the other side of the membrane. Stow reports that three to five minutes are required with his system. Even though the long equilibration time is discouraging, the approach is appealing because of the ease with which the signal from the glass electrode can be handled.

Temperature monitoring, if desired, is extremely simple to implement. The obvious sensor is a thermistor, a temperature sensitive resistor. The device would be incorporated into a bridge circuit or some other simple circuit which would produce a voltage offset as the resistance changed.

Physiologists are also interested in the total volume of air respired. Since the air is not to be collected, the total volume must be obtained by integration of the flow rate. There are numerous flow rate transducers available from the field of process control, but virtually all of them impede the flow. Such devices may sense the pressure drop through a venturi or use the flowing stream to turn a turbine. A suitable flow rate device for this application must not impede the flow; the individual under test must not be exhaling against a resisting force. There is at least one method which meets this limitation: the hot wire anemometer (24).

A hot wire anemometer is nothing more than a thin wire placed in the flow stream and electrically heated. The wire is cooled by the fluid flowing past it and, if the composition (and therefore, thermal conductivity) of the stream remains constant, then the rate of cooling is governed by the flow rate. The effective resistance of the wire is temperature dependent, and this property provides an easily measured electrical parameter whose value is governed by the flow rate. The theory of hot wire anemometery is discussed in two excellent bulletins from Thermo-Systems, Inc. (22, 23).

The most common circuit configuration is a bridge with the hot wire in one leg. A constant current is forced through the bridge, which holds the hot wire at a constant temperature. The voltage required to force the current through the bridge is monitored as an indication of flow rate past the hot wire. For the low power instrumentation used in a portable system, the recommended probe is a 0.15 mil tungsten wire (12). Signal conditioning is somewhat of a problem since the voltage is proportional to the fourth root of the flow rate. This has an advantage however, in that wide variations in flow rate can be monitored over a limited voltage range.

A suitable arrangement for the physical placement of the sensors is quite easily devised. The sensors are placed in the "out" stream of a diaphragm mouthpiece similar to that used by scuba divers. Since the chemical transducers have response times of around a second, they will not respond much to the air flowing rapidly past them during the exhale action, but rather will analyze the air trapped in their vicinity while the individual is inhaling. This is exactly the response desired, since the trapped air will be the last air exhaled and will be alveolar air from the depths of the lungs. A mouthpiece with oxygen and carbon dioxide electrodes fitted to it is available commercially from Chemtronics, P. O. Box 6996, San Antonio, Texas 78209.

OPERATION OF THE OXYGEN ELECTRODE

Although several types of oxygen electrodes have been developed, using various electrochemical cells and membranes, all produce small currents proportional to the partial pressure of oxygen at the surface of the electrode.

The operation of the galvanic oxygen electrode was thoroughly investigated by Mancy and his co-workers (14) in 1961. These researchers used a small cell composed of a lead anode and a silver cathode with an electrolyte of KOH. The cell was covered with a membrane of polyethylene, as illustrated in Figure 1.

The cell is virtually short circuited through a microammeter or other current measuring device. With these electrode materials and electrolyte, oxygen can be reduced spontaneously at the silver cathode.

$$O_2 + 2H_2O + 4(e) \rightleftarrows 4OH^-$$

Oxygen is depleted rapidly at the surface of the small cathode, and new oxygen diffuses into the electrode system through the polyethylene membrane. The diffusion is slow and limits the current produced by the cell. The rate of diffusion is determined by the partial pressure of oxygen at the exposed surface of the membrane. Mancy hypothesizes that the familiar diffusion layer is set up principally

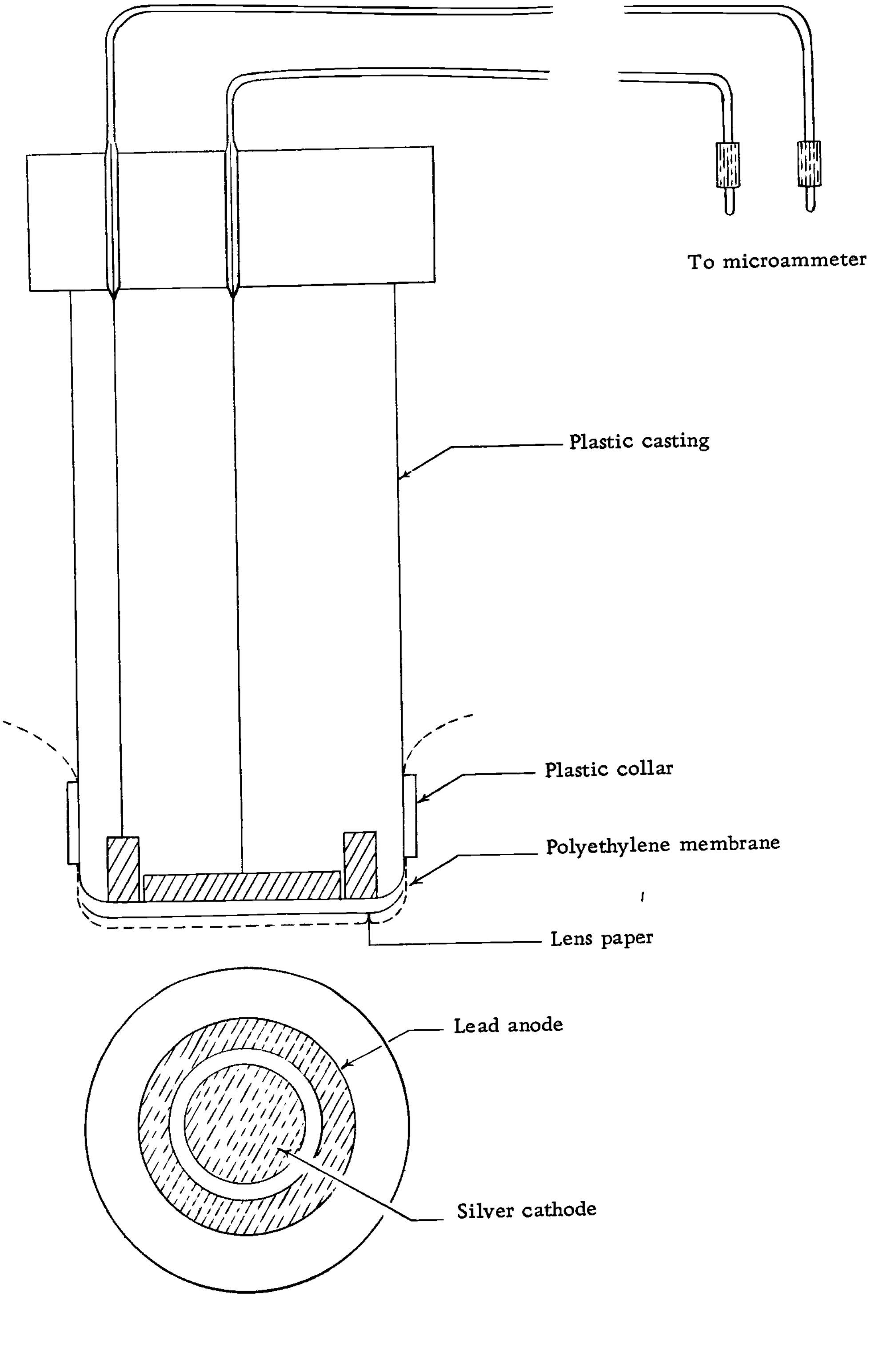


Figure 1. The galvanic cell oxygen analyzer.

in the membrane itself, with relatively rapid diffusion in the electrolyte.

Other workers have investigated similar cells using other electrode systems. Neville (16) developed a cell using a gold cathode and a cadmium anode in an electrolyte of saturated KCl. This cell was also galvanic; oxygen was spontaneously reduced at the cathode. Kinsey and Bottomley (10) used a gold anode and a silver cathode. In this system it is necessary to apply a potential to the cell to cause the reduction of oxygen. The polarographic plateau for the reduction of O_2 to OH^- extends from about 0.5 to 1.2 volts.

The Beckman oxygen electrode was used in this work. Beckman's electrode uses a silver anode and a gold cathode with KCl electrolyte. When a potential of nominally 0.8V is applied to the cell, oxygen is reduced at the gold electrode and the silver is oxidized to AgCl. The currents produced are on the order of 10^{-6} amp, so the cell may be operated for long periods of time without appreciable alteration of the chemical system. Eventually, a layer of silver oxide forms at the anode and the hydroxide level builds up in the electrolyte. The cell must then be cleaned and recharged with fresh electrolyte.

A number of membrane materials have been investigated for use with oxygen electrodes. Kinsey discusses characteristics of several of these in his paper. Polyethylene, polypropylene, and

Teflon have been the usual materials used, with thicknesses ranging from 0.5 to several mils. Thin films allow relatively large currents to flow, because oxygen can diffuse into the cell rapidly. Thicker films transport oxygen more slowly and produce lower currents. The lower currents are desirable in most applications because the cell can last longer between servicing.

The temperature coefficient of the electrode is highly dependent on the film used. As the temperature rises, the diffusion coefficient of the oxygen in the membrane increases. The result is an increase of diffusion current of about 5%/°C. Thin films give somewhat lower temperature coefficients; thick films somewhat larger. Kinsey found the temperature coefficient of one mil polypropylene to be 4%/°C. and that of one mil polyethylene to be 5.3%/°C. Rayment (18) also studied Teflon and found a coefficient of 8.6%/°C. Although there is some disagreement between these authors on the actual values of these coefficients (no doubt due to variations among the particular membranes used), it seems clear that Teflon has a much larger temperature coefficient than either of the other two materials. Nevertheless, Teflon is generally chosen as the membrane material because its other properties, mechanical stability and chemical inertness, outweigh the disadvantage of a large temperature coefficient. The electrode used in this work was covered with a Teflon membrane supplied by Beckman with the electrode.

The variation in the response of the electrode with temperature may be compensated in two ways. In experimental work it is easy to use a nomograph to adjust the output current of the electrode at various temperatures to a corresponding current at a standard temperature. In automatic applications or in field work, this is not always practical. Beckman has built into their electrode a thermistor with a negative temperature coefficient approximating that of the membrane. In the commercial oxygen meter, this thermistor is used to automatically compensate for temperature change in the range of 15 to 45°C. Briggs and Viney (1) have studied methods of compensation employing the thermistor in a bridge circuit. Unfortunately, the temperature coefficient changes each time the membrane is changed, probably because the membrane is stretched differently each time, so no one compensation can be used universally.

Because temperature compensation must be changed each time the electrode is recharged, automatic electronic temperature compensation was ruled out. Physiologists have found (3) that the temperature of exhaled air is quite well regulated by the body to 35.5°C. Since the electrode will always be operated at this temperature, compensation is not necessary. If temperature is being monitored then of course a graphical correction for variations could be employed.

Selection of the chemical and physical transducers is only part of the solution to the problem. The outputs from the transducers must be converted into some kind of usable information. Moreover, this information must appear not at the transducer, which is part of a piece of portable equipment, but rather at some fixed and remote receiving station. An electronic instrument must be constructed which will convert the transducer data into usable form and transfer these data to the receiving station for recording.

In the following pages, a survey of the methods by which this can be accomplished is presented. The method referred to as "pulse duration telemetry" was chosen for the breath analyzer. Following the survey is a detailed description of the actual methods and circuits implemented.

METHODS OF TELEMETRY

Introduction

Telemetry is a system whereby data are transferred over some distance from one location to another. The simplest example of a telemetry system is the conversion of information to a voltage level which is then measured at a location remote from the converting device (2). The connection between the voltage producing device and the voltage measuring device is a pair of wires, as shown in Figure 2. If the data must be carried a large distance before being measured, the voltage drop across the wires produces unacceptable errors in such a system.

A slight refinement of the telemetry system can overcome errors due to voltage drops. Instead of converting the information into a voltage level, it can be converted into a current, which travels over the wire to the measuring device. Since the current in a loop must be everywhere constant, there are no losses over distance. Such a system is illustrated in Figure 3.

Still, there are disadvantages to such a system. If information about a number of variables is needed concurrently, then a large number of wires must be used, as each voltage or current level requires at least one wire, even if a common return is used. A

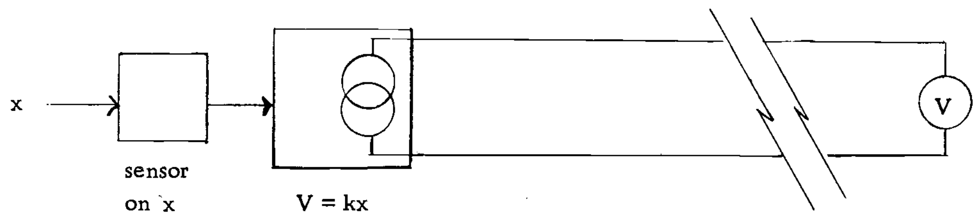


Figure 2. Wired telemetry by remote measuring of voltage.

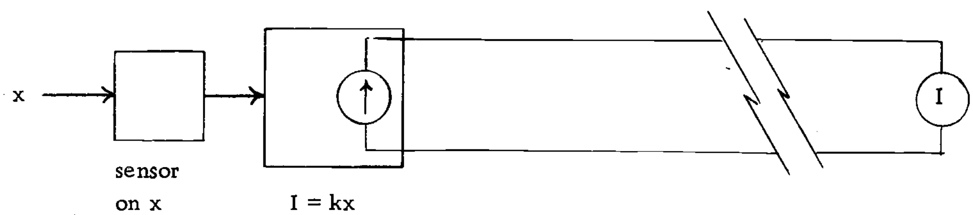


Figure 3. Wired telemetry by remote measuring of current.

superior system may be devised by using one wire to carry many different kinds of information. This is called multiplexing.

If the variables are not changing too rapidly, then a time sharing form of multiplexing may be used. Each variable is sent over the wire for only a short period of time. Then another variable is sampled; then a third; and so on, until all the variables have been covered. After the last variable has been sent along the wire, the first one is sent again, and the process continues in a cyclic manner. The repetition rate can be quite high, as it must be if some of the variables are changing rapidly or if there are a large number of variables which must be transmitted over the same wire.

As the repetition rate increases, however, new problems arise. The inductance and capacitance inherent in a long wire make rapid changes of voltages or currents impossible. The information must be converted into some other form, which is not sensitive to these problems. This may be accomplished by changing from a voltage or a current analog into some time-based system. Several possibilities are open to the designer, including pulse duration, variable frequency, or a pure digital system. These will each be discussed in detail.

But first, let us remove an encumberance which, for purposes of generalization, we have carried this far into the discussion; the wires. Although a time analog can be carried over a wire with better fidelity than a voltage or current analog, the problems discussed above will still exist at high rates of repetition. Moreover, it is not always feasible to use wired connections, as, for example, in systems in which the sending and receiving devices are mobile with respect to one another, or at very large distances from one another, or on opposite sides of some barrier through which it is impossible or undesirable to pass a wire. In cases such as these, the medium of transmission is a radio wave rather than a wire. The information needed, converted to a frequency, pulse duration, or digital code, is impressed upon a radio carrier and transmitted to the receiving station.

The use of radio waves for transmission of information creates the possibility of another form of multiplexing in addition to the time-sharing introduced above. In radio transmission, information is impressed upon a carrier of some frequency. It is quite feasible to impress the information on some subcarrier, which is in turn impressed on the radio carrier. A number of these subcarriers may be put on the radio carrier at once, as long as the subcarriers differ sufficiently in frequency. This type of multiplexing is called frequency multiplexing.

The use of a radio carrier also introduces new difficulties. The behavior of the receiving and transmitting equipment must be compatible with the method of presenting and multiplexing the data.

An excellent analysis of the use of radio for telemetry appears in the text by Nichols and Ranch (17).

Signal Conditioning

For a single channel of data, a simple method of transmission which is not subject to the inherent errors of a voltage or current level is a frequency analog. The parameter of interest is converted into some frequency. As the parameter changes, the frequency of the transmission changes. The simplest method of implementing such a system is to place an active element in the frequency determining circuitry of an oscillator. For example, Figure 4 shows how temperature may be converted to a frequency by placing a thermistor in the RC network of an oscillator (13). Other possibilities include the use of a photocell as R to measure light intensity, and the use of a strain gauge as R to measure acceleration or stress. Figure 5 illustrates the use of a variable inductor in an LC oscillator circuit.

Often a variable impedance transducer is not available for a particular application, but instead a voltage producing transducer may be used. This is particularly true in chemical systems, where the designer makes use of glass electrodes or electrochemical cells. In such cases a voltage to frequency converter can be used. Typically this will be some kind of blocking oscillator. The voltage from the transducer will be used to charge a capacitor to some preset level.

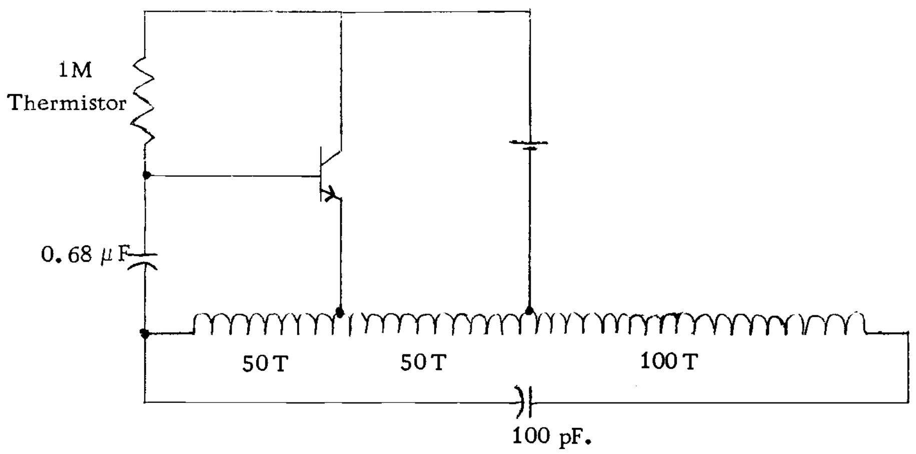


Figure 4. Variable frequency oscillator dependent on temperature.

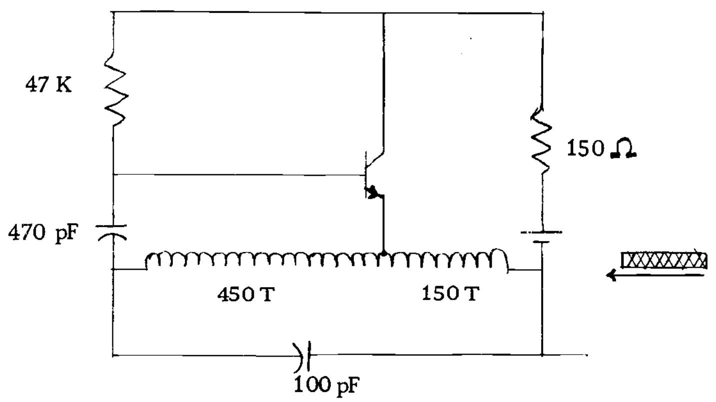


Figure 5. Oscillator with variable inductor.

When this level is reached, the capacitor is discharged and charging begins again. The resulting potential across the capacitor is a saw-tooth wave whose frequency is related to the charging and discharging currents to the capacitor.

Figure 6 illustrates a simplified voltage to frequency converter. A current equal to e(in)/R flows from the capacitor and a charge builds up across it. This charge is manifested as a voltage level e(o). The capacitor is discharged by some kind of switch when e(o) reaches a preset level, sensed by the level detector.

The accuracy of such a system depends on the ideality of the charging circuit and on the stability of the level detector in the discharge circuit. With the advent of precision DC amplifiers, error in the charging circuit can be well controlled. The amplifier must have a low drift in voltage offset over a long period of time. Present monolithic operational amplifiers easily meet this criterion.

Other than the amplifier, the critical elements in the ramp generator are the resistor and the capacitor. The resistor should be of a type with a low temperature coefficient. The capacitor is generally a polystyrene type, as these exhibit very low leakage. Obviously, current leaking across the capacitor is lost charge and results in an error in the output. Leakage in capacitors is directly proportional to the capacitance; consequently the capacitor should be of as low a value as practical, generally not over one microfarad (7).

Some method must be devised for detecting the voltage at the output of the ramp generator and discharging the capacitor when a particular level is reached. The level detector and the discharge circuit are not necessarily distinguishable from one another.

Figure 7 shows a circuit in which the discharge is accomplished by means of a unijunction transistor (20). The emitter junction becomes forward biased when $V_e = \eta(V_{B1} - V_{B2}) + V_d$ where η is the intrinsic stand off ratio (about 0.6), V_{B1} and V_{B2} are the voltages at the upper and lower base connections, and V_d is the potential drop across the junction (about 0.6V). In the circuit shown, B2 is about at e(o) because R3 is quite small. The emitter is connected to the summing point and must be at virtual ground. Substituting these values into the equation for the firing potential of the unijunction, the condition for conduction is

$$e(o) \leq -1.5(V_{B1} + 1).$$

When e(o) reaches this potential, the unijunction turns on and the capacitor is discharged rapidly, until the potential across it falls below V_d. The result is a periodic discharge of the capacitor and a sawtooth waveform.

There are a number of serious problems associated with this kind of a circuit. Foremost is the drift of unijunction firing potential with temperature. This is primarily an effect on the value of the

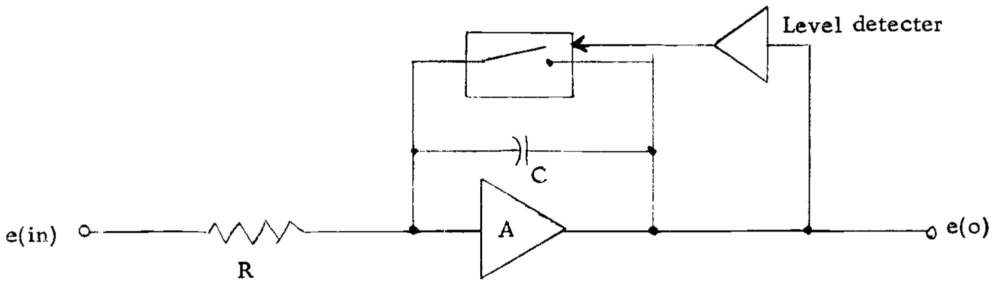


Figure 6. Sawtooth Wave Generator. The capacitor is charged by current e(in)/R. When a preset voltage level is reached at the output of the amplifier, the level detector discharges the capacitor.

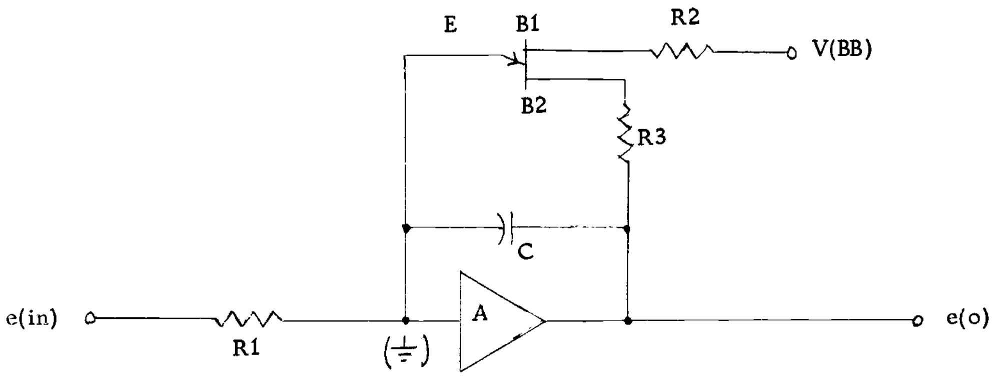


Figure 7. Sawtooth generator with frequency proportional to input voltage. The unijunction acts as both the level detector and the discharging switch.

intrinsic standoff ratio. An uncompensated unijunction relaxation oscillator may drift in frequency as much as 5% over a 20° temperature range. For a full discussion of this problem and its remedy, see the GE Transistor Manual (6).

An associated difficulty is variation in the turn-off potential of the transistor. The capacitor may not always discharge to the same level. Either kind of error produces unacceptable drift in the frequency of oscillation.

A more successful circuit can be devised by sacrificing some of the simplicity of the previous circuit. Instead of using a single semiconductor element to sense a voltage level at which discharge is to occur, as in the unijunction circuit, a precise voltage comparator can be constructed which changes state in some way when the output voltage of the ramp generator reaches a preset level. A high gain differential amplifier will perform such a function.

The base of one of a pair of matched transistors is biased with a reference current, and the other member of the pair is biased from the output voltage of the integrator. The two matched transistors are connected in a common emitter configuration into amplifying transistors such that the final stage saturates either on or off, depending on the relative levels of the two input voltages. Typically, the comparator may be a differential operational amplifier with no feedback, so that the output rises rapidly from one saturation limit to the other when the input voltages cross. For high speed operation, the amplifier must be protected against true saturation at either the inputs or the output. This may be done with clamping diodes. A better solution is to use an amplifier designed for comparison purposes such as the Fairchild μA710.

The output of the comparator can be used to close a reed relay across the capacitor, as shown in Figure 8a. The capacitor will rapidly discharge to zero when the relay closes. If a high rate of operation is necessary, a mechanical switch may not be satisfactory. It might be replaced by a transistor, but this reintroduces the problem of indeterminate voltage level at the end of the discharge. A field effect transistor, connected as shown in Figure 8b, should operate successfully.

A voltage to frequency converter operating within well defined limits can be constructed using two comparators, as shown in Figure 9. A ramp is generated from the input voltage by the integrator. When the ramp reaches a preset level, V(ref), comparator A2 turns on a flip-flop which feeds a voltage, through a small resistor, into the summing point of the amplifier. The current into the summing point from this source is much larger than, and of opposite polarity to, the current due to the input voltage. Consequently, the rate of discharge is very rapid compared to the charging rate. The only real limiting factor is the slewing rate of the

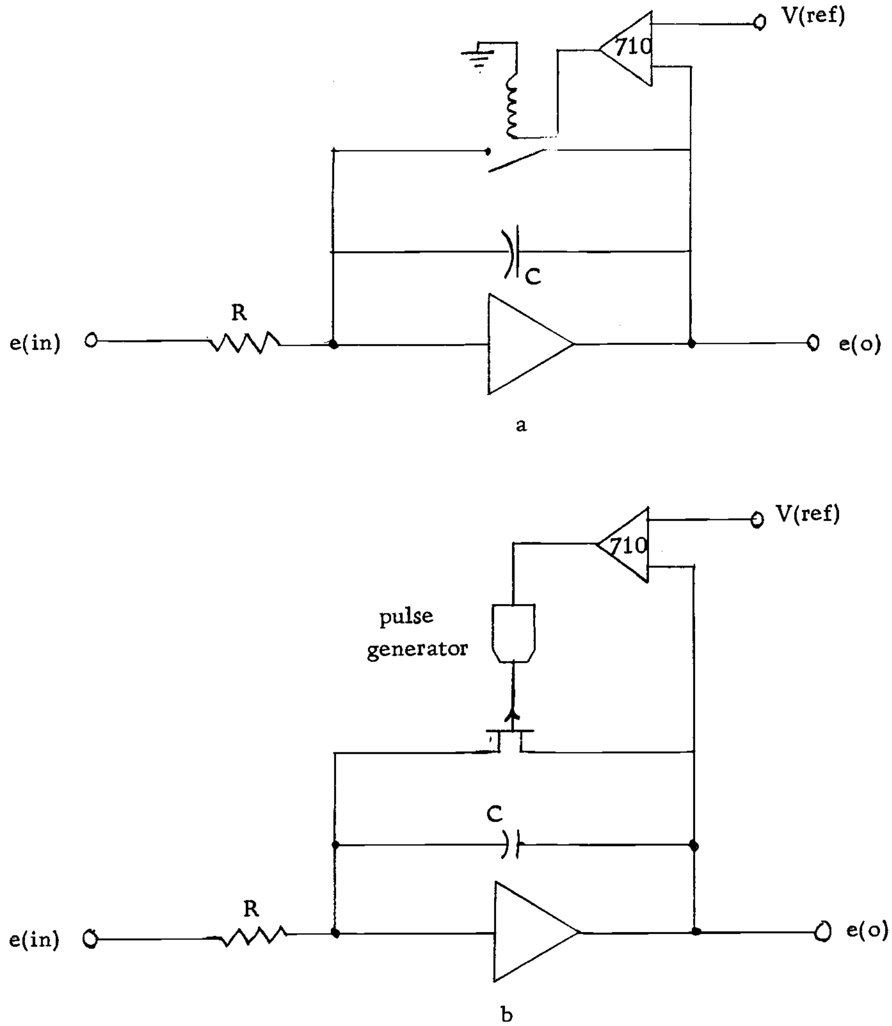


Figure 8. (a) Integrating capacitor discharged by reed relay.
(b) Integrating capacitor discharged by FET.

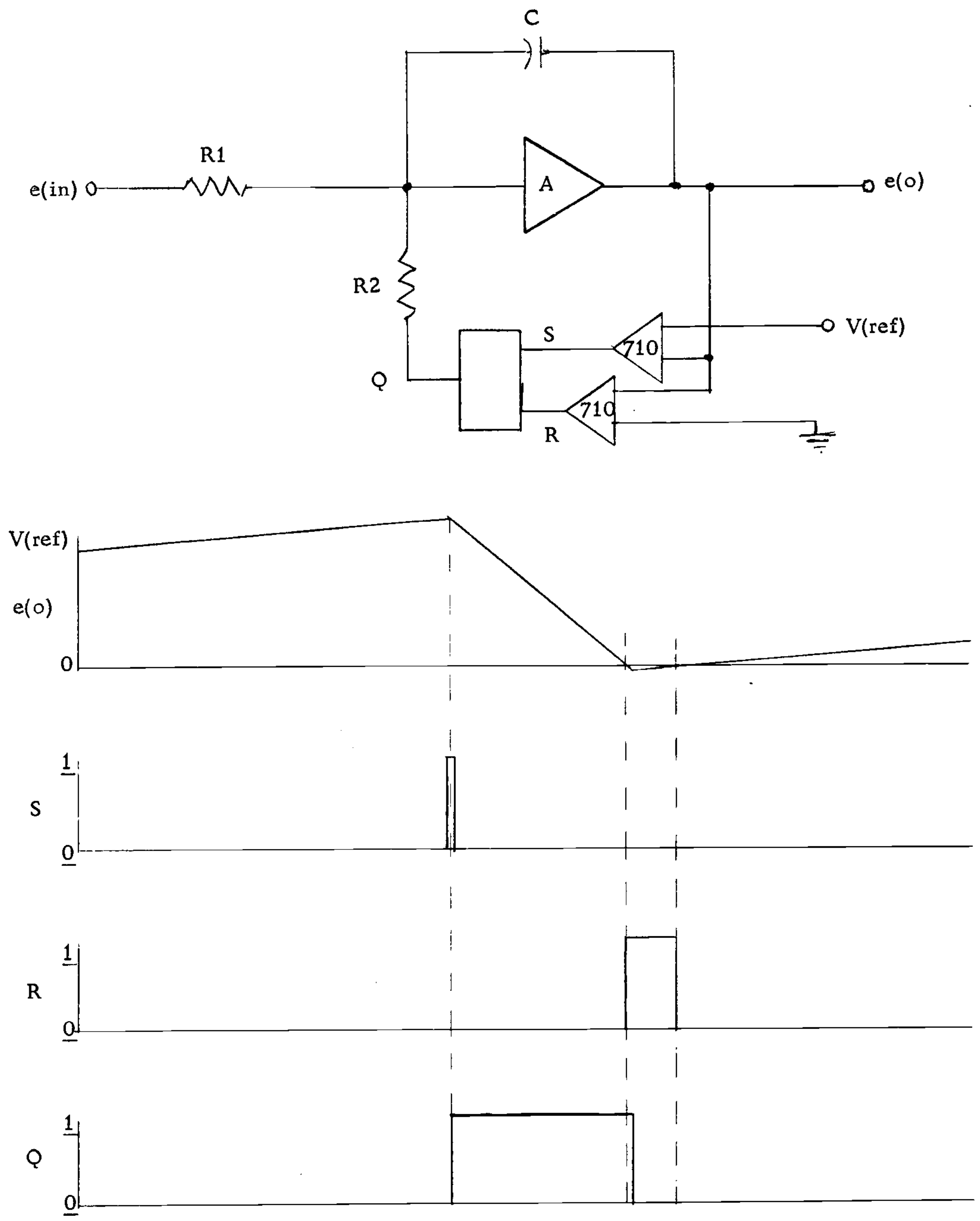


Figure 9. A controlled sawtooth generator with timing diagram. The slight negative swing in the output is due to time delays in the switching networks. The time scale is large; discharge time is on the order of a microsecond.

amplifier. A second comparator, A3, is used as a zero crossing detector. It serves to turn off the flip-flop and, hence, the discharge current into the summing point when the output of the integrator has reached zero. Charging may then begin again.

Such a circuit is considerably more complex than the simple circuits with which we began this discussion, but it is far less sensitive to drift than the simpler circuits. Precision has been increased by minimizing the number of truly analog elements and introducing instead some devices which must only say "yes" or "no," not "how much."

There are many variations of these variable frequency sawtooth generators, but most rely upon the principles which have been illustrated above. For further discussion the reader is referred to the Philbrick Operational Amplifier Manual (7), paragraphs II. 41, II. 49, III. 16, and to the GE Transistor Manual chapter referred to earlier.

The sawtooth generators discussed above convert the voltage of interest into a period of time; namely, the period of the sawtooth wave. It is a simple step from such a circuit to a pulse duration circuit. Once again a period of time is the parameter of interest. The pulse duration circuit is especially amenable to time multiplexing.

In a pulse duration operation, there are only two possible states: either a voltage is present or it is not. The length of time during which the voltage is present is proportional to the input voltage which is to be measured. In practice, the circuit consists simply of a ramp generator and a comparator to compare the ramp voltage with the unknown voltage. When the ramp exceeds the unknown voltage, the comparator changes state.

The ramp generator is exactly like the integrator discussed under the previous section, except that the charging current is derived from a reference voltage and the discharge pulse is derived from a clock. The system is illustrated in Figure 10.

The sequence of operation may be described as follows. Initially, the output of the amplifier is zero. The current from the reference voltage source into the summing point charges the capacitor and the voltage at the output of the amplifier rises linearly with time. The output of the comparator is a logical "1." At some point along the ramp, the ramp voltage equals the voltage to be measured. When the ramp exceeds the unknown voltage, the output of the comparator shifts suddenly to a logical "0." The ramp continues to rise until a clock pulse occurs. The clock triggers a flip-flop which discharges the capacitor in the same manner as in the sawtooth generator discussed above. As soon as the integrator output falls to zero, the comparator once more produces a logical "1." The result is a train of pulses in which each pulse is initiated by the clock and ends after some time interval which is proportional to the voltage to be

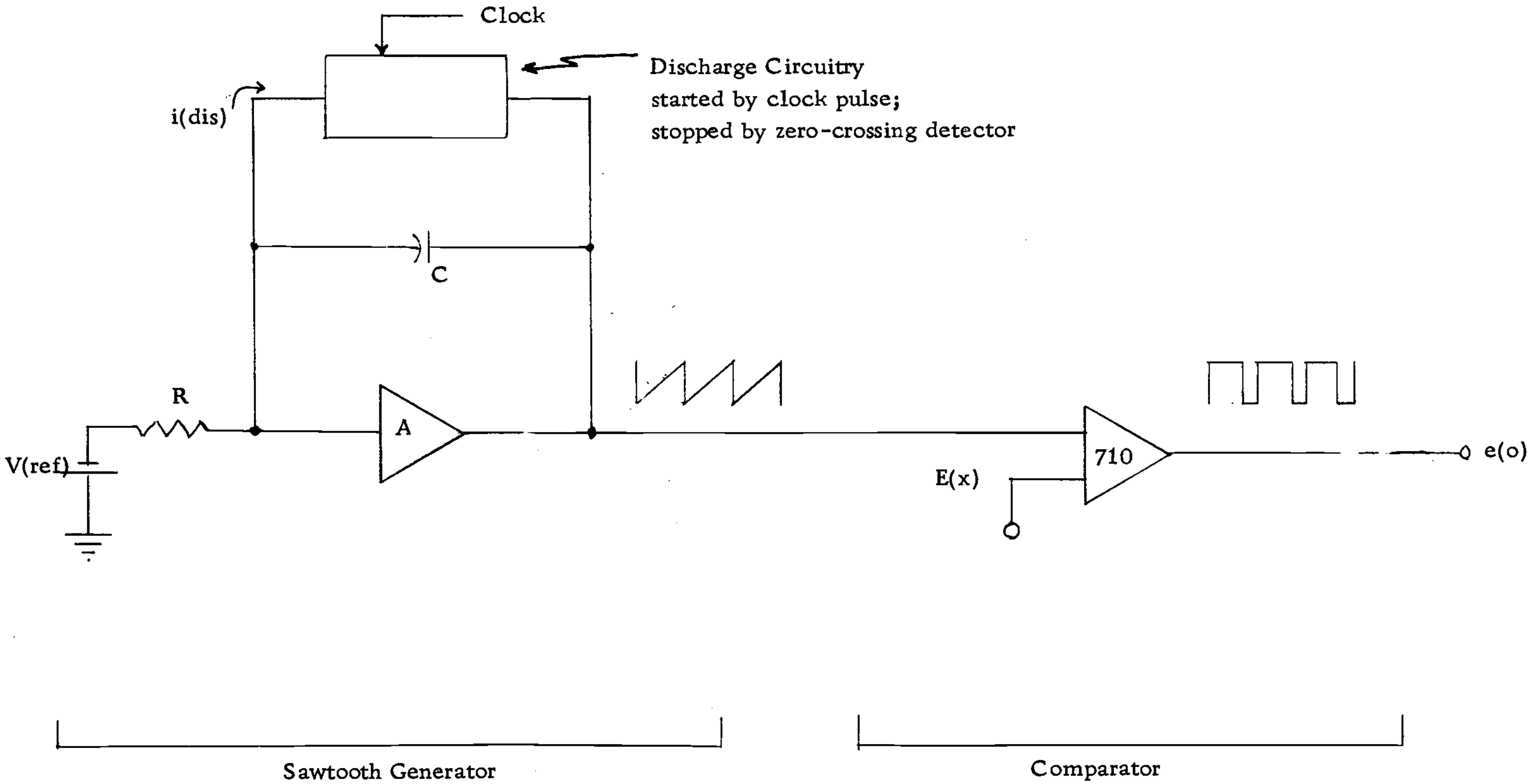


Figure 10. Pulse duration system. The width of the pulses is proportional to E(x).

measured.

From pulse width telemetry it is once again a small step to the next type: pulse code or true digital. In the pulse width measurement, the quantity transmitted is still analog, namely time. In the pure digital system, we release our last hold on analog quantities and the only importance of time is to provide a sequential medium for the transmission.

The first chore in changing from pulse width to pulse code is to change slightly the character of the ramp. In the previous systems, the ramp was generated by integrating a current with respect to time. In the digital system a staircase rather than a linear sweep is required. The method used for generating this staircase may seem somewhat cumbersome at first, but its utility will be obvious later.

A clock is needed to trigger each step in the staircase. The accuracy of the clock frequency is not important, it is only necessary that it operate at high speed. The clock pulses are counted by binaries set up to count in some particular code. For purposes of simplicity, we shall assume the use of a conventional 2^n BCD code (binary outputs weighted 1, 2, 4, 8, etc.). The outputs of these binaries are used to turn on and off current drivers which produce constant currents whose magnitudes are weighted 1, 2, 4, etc., in accordance with the particular binary output with which they are switched. The switches may be either solid state or reed relays.

These currents are all summed and dropped across a precision resistor. The voltage drop across this resistor will increase by one unit each time the counter counts a clock pulse. This is the required staircase. A significant point is that we have not only generated a staircase, but we have available simultaneously a binary number for each level of the staircase. Figure 11 illustrates one way in which the staircase can be generated (5, 19).

Figure 12 shows how the staircase is implemented into the complete system. A comparator is used as in the pulse duration system to detect the point at which the staircase first exceeds the voltage to be measured. As soon as the comparator indicates that this condition has been fulfilled, the binary number in the counters is transferred to the data inputs of a shift register. This number represents, within an accuracy of one bit, the level of the unknown voltage. (The term "bit" here signifies the magnitude of each step in the staircase function.) Upon some sort of signal, either internally generated or coming in from outside, the data in the shift register is shifted out at a constant rate.

The end result is that a binary number has been obtained which represents the level of the unknown voltage, and that number has been sent onto a transmission line in a series of pulses. The shift register has served to give us a serial transmission of data which was entered into it at one time, or in "parallel."

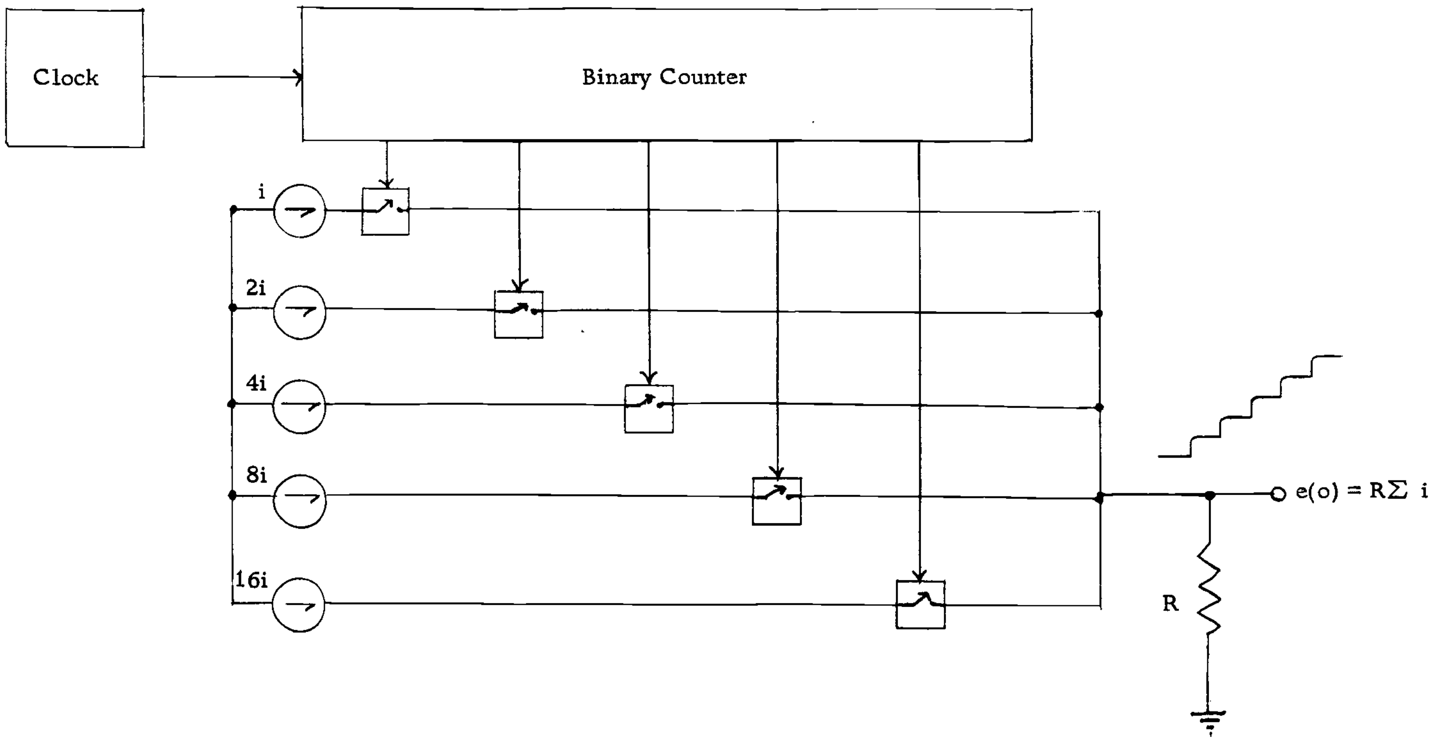


Figure 11. Generation of staircase voltage by use of binary counter and current sources.

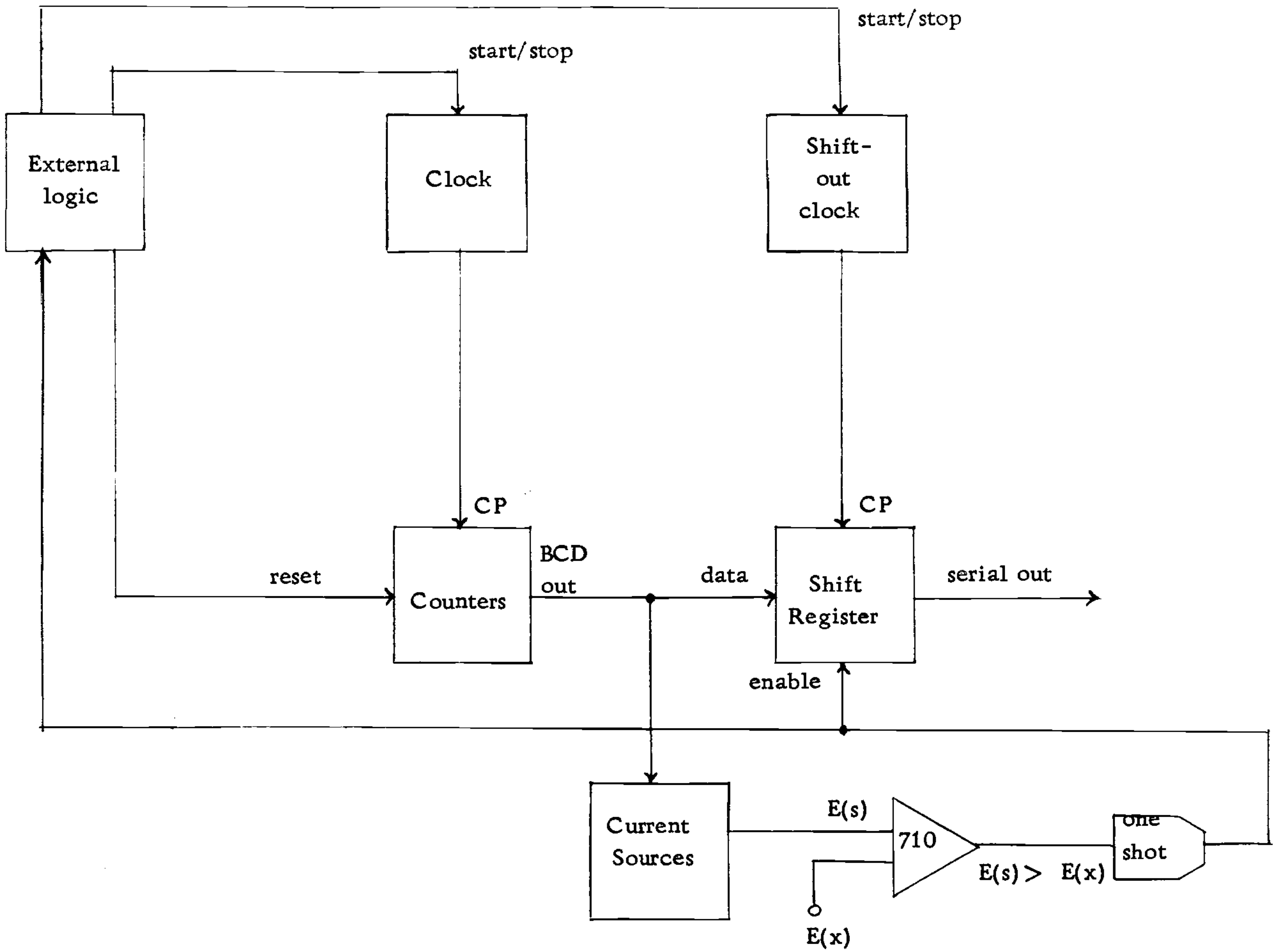


Figure 12. Block diagram of pulse code system.

Recovery of the Data

Several methods have been presented for converting information into a form compatible with transmission. At the receiving station, the transmitted form of the information must be reconverted into some kind of display.

Pulse duration signals can be converted into a voltage analog for recording by running them through a low pass filter. The voltage at the output of the low pass filter will be related to the average "on" time of the pulses. A variable frequency signal can be converted to a voltage analog in a similar manner. Each cycle of the wave train triggers a short pulse from a monostable multivibrator. These pulses are then "averaged" by a low pass filter and fed to a recorder.

The digital transmission from pulse code modulation is readily converted to a digital display. The train of pulses is fed sequentially into a shift register. At the end of the series of pulses, the register contains the same binary number that was on the staircase generator in the transmitter when transmission occurred. The binary number can be decoded into a decimal number, or the binary information can be changed to an analog voltage level by current summing in the same manner as used in the transmitter to generate the staircase function (19).

The preceeding discussion has presented in a cursory manner the several common forms in which data can be handled in a telemetry system. There has been no attempt to delve into the detailed circuitry involved; of course any electronic circuit turns out to be more complex in practice than it does in principle. There exists a multitude of variations on each of the basic methods presented, but the principle of operation of these variations is usually the same as what has been outlined above.

THE SYSTEM USED

Description

In the initial work on the breath analyzer (20), a voltage controlled oscillator was used, with simple FM telemetry of the type discussed previously as variable frequency telemetry. Such methods are frequently used to transmit biological data (13, 25). Because of problems with random drift of frequency, temperature dependent drift of frequency, and because of the necessity of some kind of multiplexing, the variable frequency system was discarded in favor of a pulse width system. Pulse code was deemed excessively complex and expensive.

A block diagram of the pulse width system is shown in Figure 13. The signal conditioners represent the interface between the electrochemical transducers and the telemetry system. The telemetry system was designed to operate on inputs varying from zero to two volts, so the signal conditioners transform the output of the transducers into voltage levels within that range.

The pulse generators are simply comparators which compare the output of the signal conditioners with a ramp voltage as described previously under the discussion of pulse width telemetry.

The serial multiplexer consists of a "ring counter" and a series

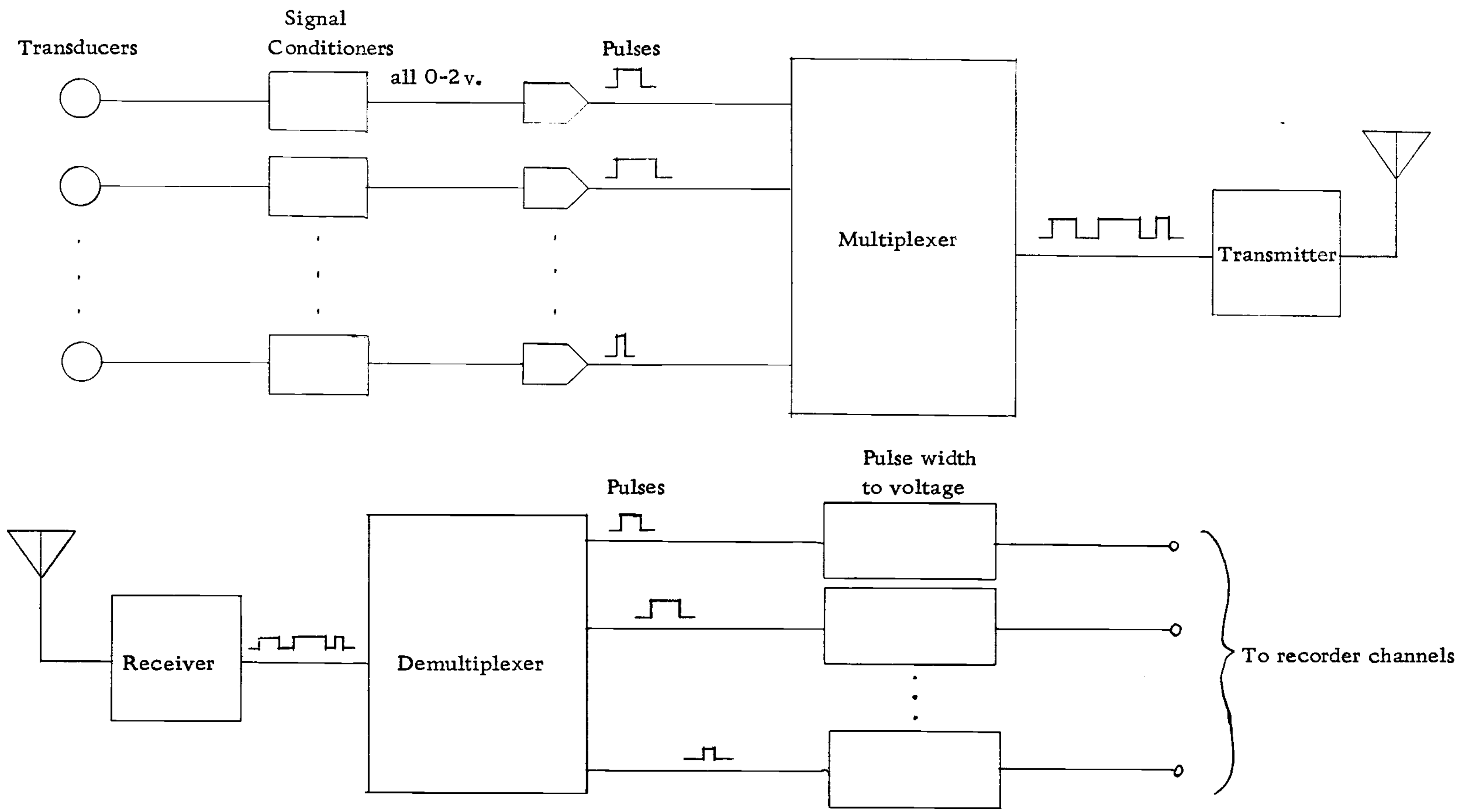


Figure 13. General scheme of telemetry.

of gates. The output of each pulse generator is connected to the input of a gate. The gates are turned on one at a time by the ring counter and their outputs are tied together. The result is a train of pulses in which the pulse from the first channel is followed by one from the second, and then by one from the third, and so on. The last pulse in the train contains special synchronizing information to identify the end of the pulse train. A pulse from the first channel follows the synchronizing information and the cycle is repeated. The multiplexer was constructed to permit up to 7 channels of information besides the synchronization.

The pulse train can be transmitted by a standard commercial FM telemetry transmitter to the receiver, where it is demodulated and recovered.

The pulse train from the receiver is demultiplexed--the pulses are separated into different channels. The synchronizing information is used to assure that a given pulse in the train always ends up in the same channel.

Finally, the pulses are converted into analog voltage levels which can be recorded on a strip chart recorder. This operation is performed by charging a capacitor at a constant rate during the time the pulse is on. The resulting voltage is proportional to the on time of the pulse and is fed to the output through a sample and hold circuit.

On the following pages the actual circuits used are presented.

The position of each circuit in the system is indicated by the block diagram on Figure 14. Each block in Figure 14 is a separate circuit wired on its own circuit board. The schematic diagram of each circuit is followed by an explanation of the operation of the circuit. The circled numbers at the bottom of each schematic indicate pin connections on the circuit board sockets.

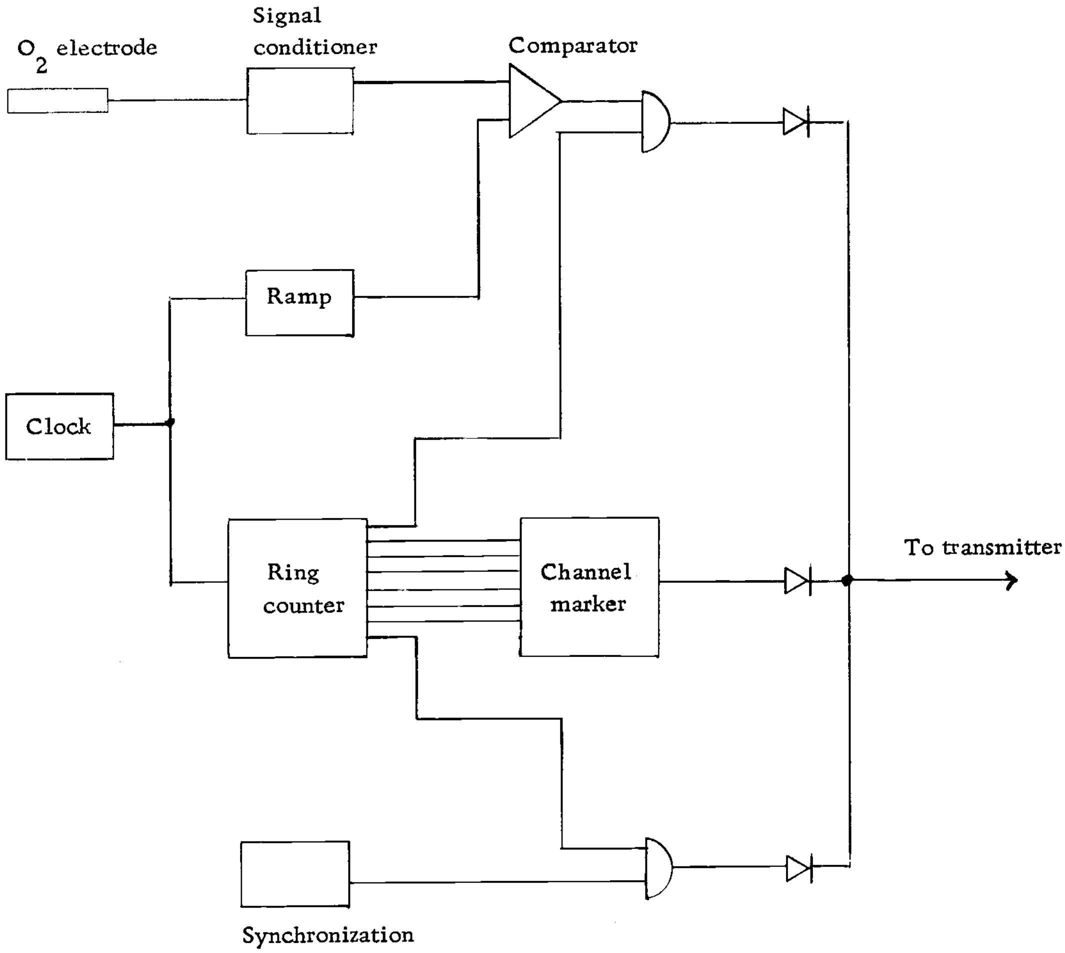


Figure 14. System used for transmission of oxygen data by PDM telemetry.

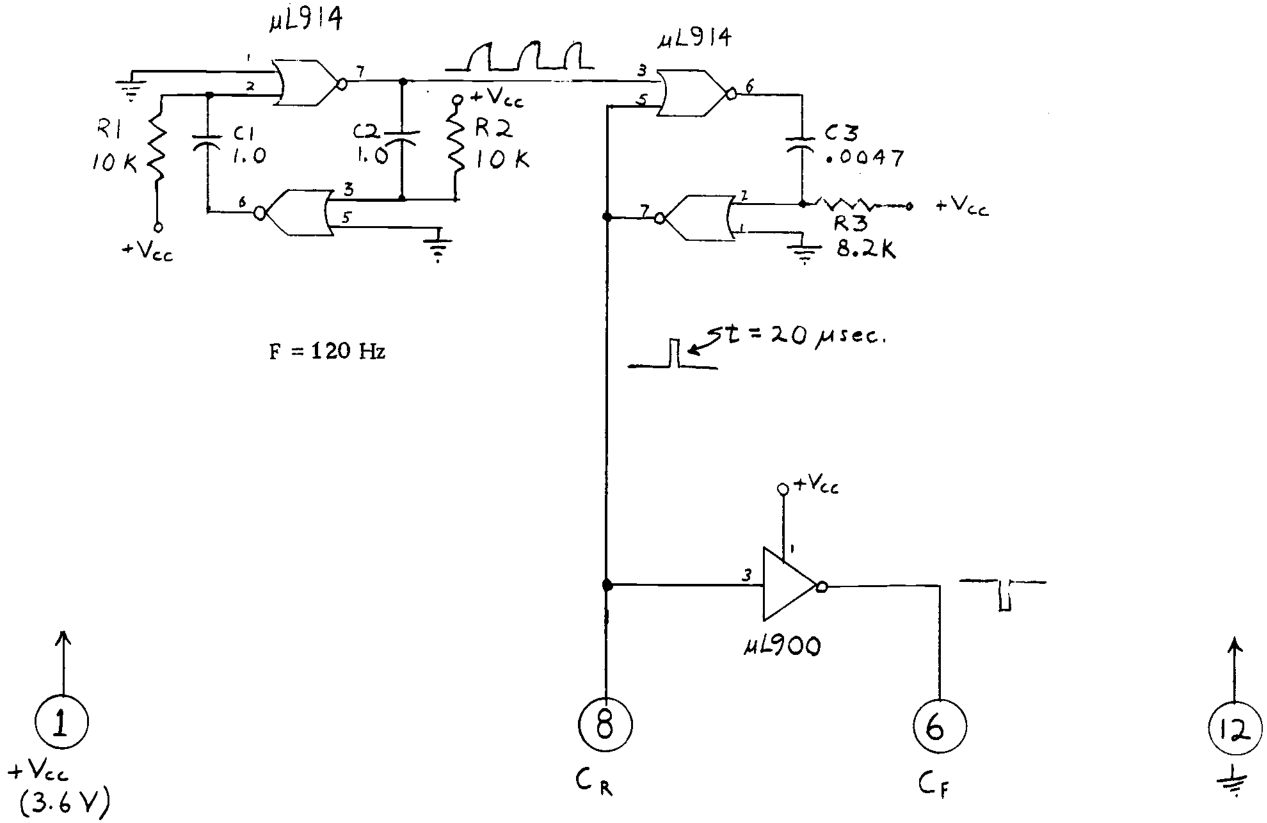


Figure 15. Clock.

Clock Circuit

The basic operating frequency of the system is controlled by a clock. The requirement for the clock circuit is that it produce a short (several microsecond) logical zero every few milliseconds.

The clock frequency is set by an astable multivibrator using two μL914 gates (11). The frequency of the multivibrator is governed by R_1C_1 and R_2C_2. These components determine the period of the two halves of the square wave. For proper operation, the two time constants must be nearly equal, so the required short 0 time and long 1 time cannot be directly generated by the astable circuit. Instead the two time constants are set equal so that a symmetric square wave is produced. The frequency is 120 Hertz.

The leading edge (0 to 1) transition of the square wave is used to trigger a monostable multivibrator which produces the required short pulse. The period of the short pulse is determined by the values of R_3 and C_3. For the components used, the monostable period is 20 microseconds. The pulse produced by the monostable is a logical 1. This line is labeled C_r, rising clock pulse. Most points in the system which must be clocked require instead a logical 0 pulse, so the output of the monostable is inverted by a μL900 buffer-inverter. The uL900 also provides a large fan-out; that is, a large number of devices may be operated from it without overloading the output. The buffer output is labeled C_f, falling clock pulse.

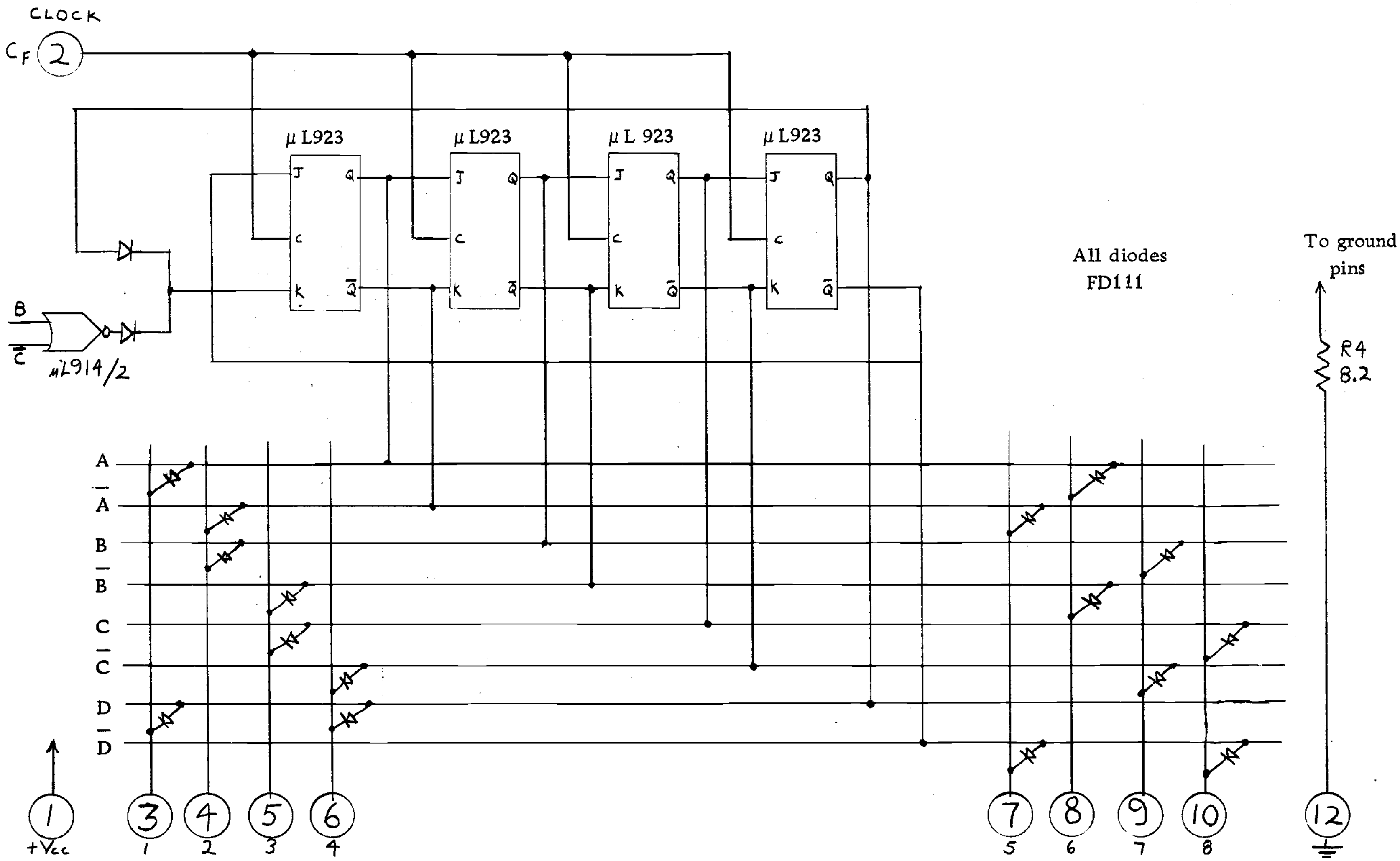


Figure 16. Ring counter.

Ring Counter

The output pulses from each channel are to be fed one at a time onto a common transmission line. A device is needed which will turn on channel one for one time period, then channel two for one time period, then channel three, and so on. This is accomplished by enabling a series of gates one at a time. The device which produces this enabling signal is the ring counter.

The ring counter must be able to turn on, one at a time, eight discrete outputs. The enable signal to the gates is a logical 0, so in each time period one of the outputs will have a 0 on it, all others will have 1's. At each clock pulse, the 0 will be transferred to the next channel. Since the last channel is followed by the first, the device is cyclic and is called a ring counter. Since eight discrete outputs are required, the device must have at least eight discrete states. This could be accomplished by counting clock pulses in ordinary BCD code and decoding the counter output into one of eight outputs. Three flip-flops would be required to produce eight states.

If a fourth flip-flop is used, however, a counting sequence can be generated which is more easily decoded into discrete states than the BCD code. The counting sequence generated by the circuit shown in Figure 16 is given below:

ABCD	
0000	f0
1000	f1
1100	f2
1110	f3
1111	f4
0111	f5
0011	f6
0001	f7

It may be seen by inspection that each state can be uniquely specified by using only two of the four variables. The first state is the only one which has a 0 in both the first and last places. Hence, if a point is connected through diodes to both the first and last flip-flops, that point will have a logical 1 on it at all times except during state f0. Similarly, only state f1 has a 0 in the second place and a 1 in the first place. A 0 may be uniquely specified during state f1 by ORing (through diodes, as before) B and $\bar{A}$. This process can be repeated for each of the six remaining states. The diode matrix in the ring counter circuit is used to combine the proper combinations of the four variables and their complements to uniquely decode all eight states.

Decoding Table for Ring Counter
(f states are logical 0's)

f0 = AD	f4 = $\bar{A}\bar{D}$
f1 = $\bar{A}B$	f5 = $A\bar{B}$
f2 = $\bar{B}C$	f6 = $B\bar{C}$
f3 = $\bar{C}D$	f7 = $C\bar{D}$

A four bit counter contains 16 states and only eight have been used. It is necessary to determine how those other eight states behave, because the counter may start up in one of them or it may, by error, fall into one during its operation. The other eight states also form a loop.

ABCD	
1010	f8
1101	f9
0110	f10
1011	f11
0101	f12
0010	f13
1001	f14
0100	f15

The presence of this loop means that if one of these states ever appears in the counting register, then all eight of the undesired states will appear and the register will not return to the desired sequence. The decoding scheme used for f0 through f7 will not work in the second loop. Consequently, some additional logic must be introduced to force the second loop to lead into the first. The normal counting sequence will be altered so that one of the states in the second loop will lead not to its ordinary successor, but rather to one of the states in the first loop.

This is the purpose of the NOR gate on the K input of flip-flop A. The simplest way to effect the transfer is to cause state f13 (0010) to go to state f7 (0001) instead of its natural successor f14 (1001). This sequence is chosen because the desired successor f7 differs from the natural successor f14 only in the first bit. All that is required is to prevent the appearance of a <u>1</u> on bit A following state

0010. This can be done by applying a 1 on the K (clear) input of flip-flop A during state 0010. The NORing of B and $\overline{C}$ will produce the desired 1 without affecting the operation of the principle loop f0 to f7. The only time in the main loop that B and C are present in a combination which can activate this logic is in state f6, and it is followed by a 0 on the first bit anyway.

The voltage drop through the decoding diodes causes the logical 1 level not to be of high enough voltage to reliably activate the logic gates to which the outputs are connected. This problem can be solved by placing a resistor between the logic devices' ground and the system ground. The result is that the logical 0 level and logical 1 level are both raised slightly. The 8.2 ohm resistor used raises the logic levels just to the point that they can be correctly interpreted by succeeding logic. This action would not be necessary if TTL logic had been used instead of RTL, because TTL operates with a greater difference between a logical 1 and a logical 0.

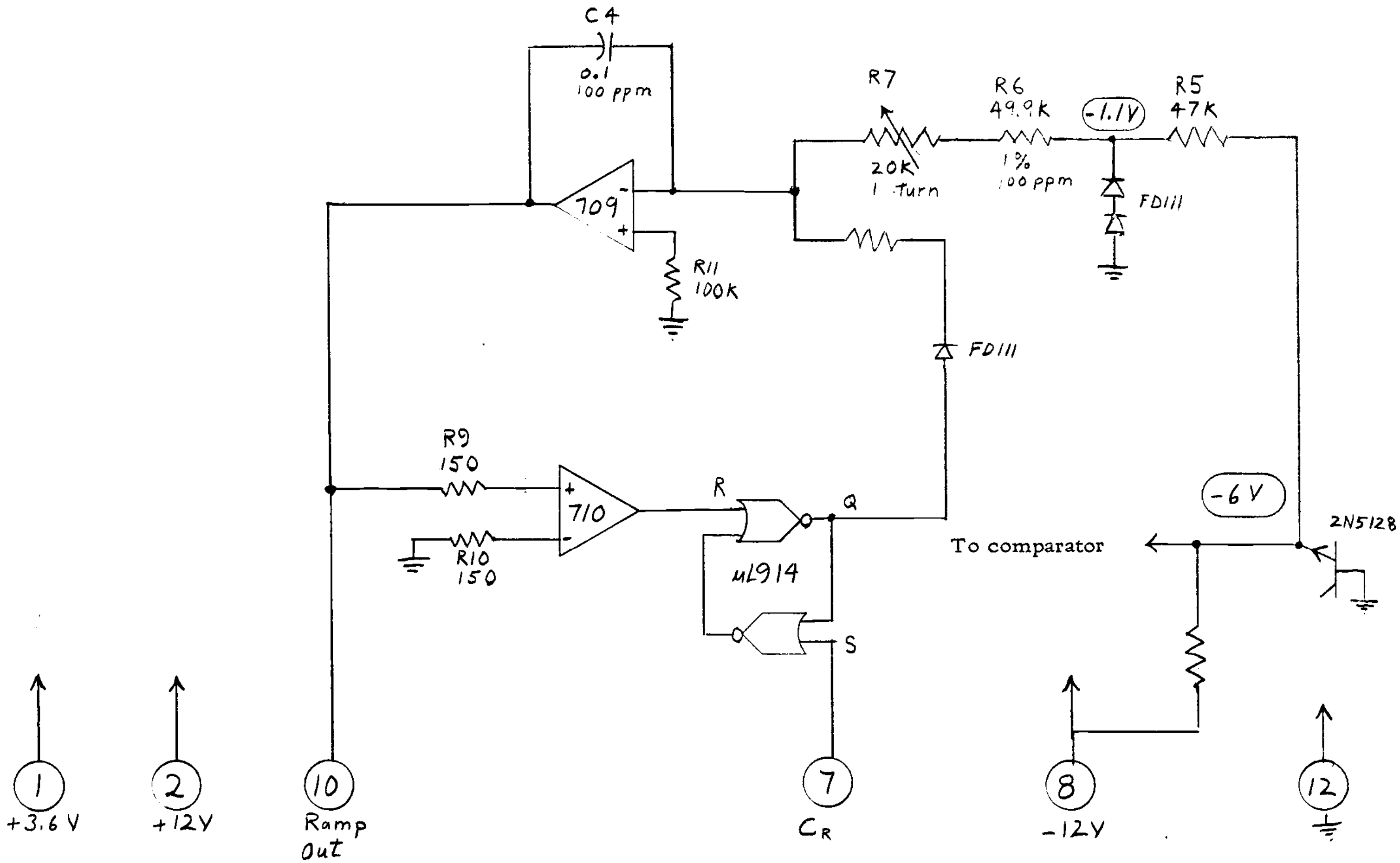


Figure 17. Ramp generator.

Ramp Generator

Since a pulse-width system is to be used, it is necessary to have a ramp voltage with which to compare the voltage from each signal conditioner. The ramp must be very linear and exactly the same each time it is produced. The method used is integration of a reference current by an operational amplifier, followed by discharge to zero volts when a clock pulse appears.

The input current to the integrator is derived from the -6 volt power supply dropped through several resistors. R5 is a current limiting resistor between the power supply and two forward biased diodes to ground. The diodes serve to clamp the R5-R6 junction at about 1.2 volts. This is the forward voltage drop for the two diodes at a current of about 100 microamps. The fixed resistor R6 and the variable resistor R7 set the current into the summing point of the integrator. The integrating capacitor is a precision polystyrene type. R11 is a 100K resistor used to make the impedance at the non-inverting input about the same as that at the inverting input. The output of the integrator is labeled Ramp Out.

The end of each integrating period is indicated by the appearance of a clock pulse. The rising clock pulse, C_r, is connected to the "set" input of an R-S flip-flop, causing the output of the flip-flop, Q, to become a 1. The flip-flop output is fed, through a diode and current limiting resistor, directly to the summing point of the integrator. The large positive current overrides the small negative current from the reference and slams the integrator output back toward zero volts. The zero volt level is detected by the μA710 comparator, which produces a pulse resetting the R-S flip-flop to Q = 0. A new charging cycle then begins.

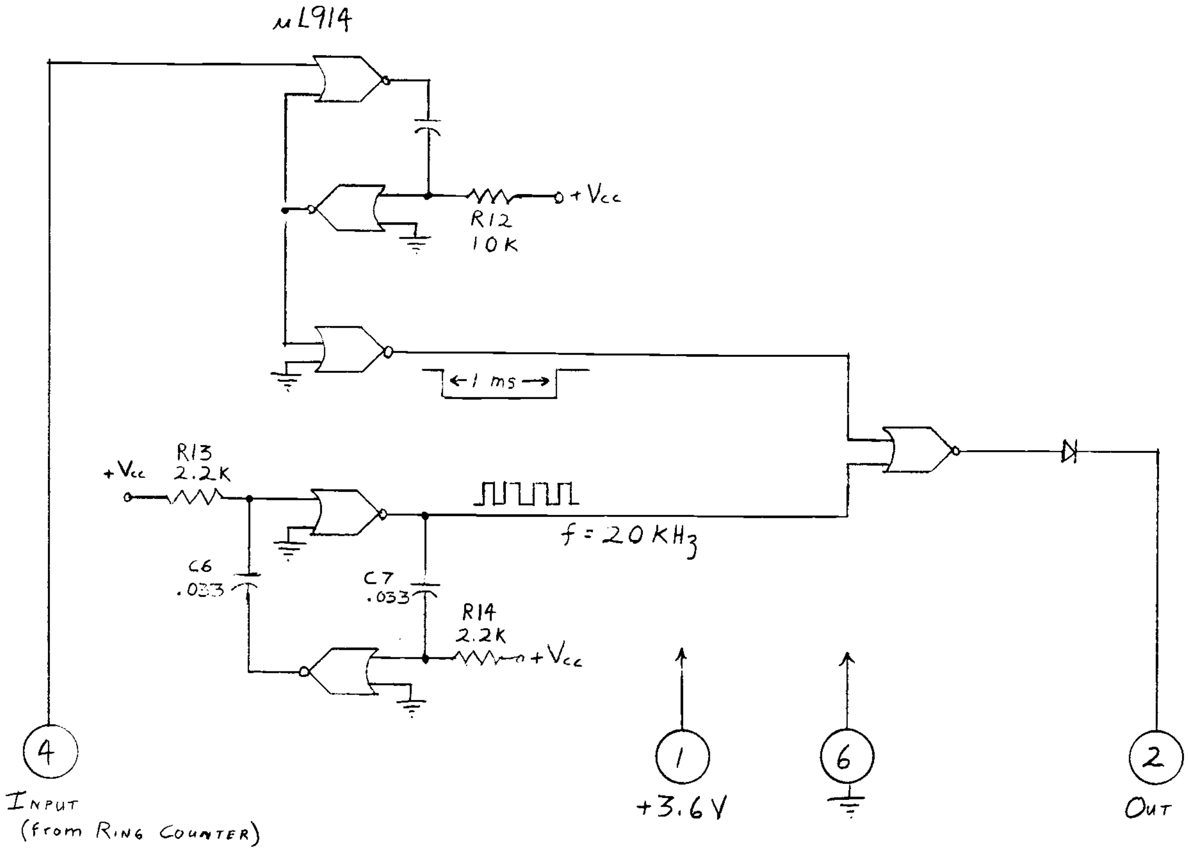


Figure 18. Synchronizing signal generator.

Synchronizing

Each channel of information transmitted will be of essentially the same form: a pulse of a particular duration. Some method must be devised for indicating which pulse came from which channel. This is accomplished by transmitting a special synchronizing pulse at the end of each cycle of eight channels. The demultiplexing equipment on the receiver will detect this signal and automatically place the next pulse on channel one, the second on channel two, etc., returning to channel one each time the synchronizing signal is received.

One channel, channel eight, is set aside for the synchronizing information. When the ring counter opens this channel, a series of 20 rapid pulses is transmitted. These can be detected at the receiving station by virtue of the fact that no other event can result in so many pulses in such a short time.

The required function is realized by the circuit on Figure 18. An astable multivibrator, similar to the one used in the clock, is designed to operate at 20 kHz. When the input from the ring counter goes to 0, indicating channel eight, a monostable opens a gate and lets through the 20 kHz. signal for one millisecond. The output of this gate is tied onto the main transmission line with a diode.

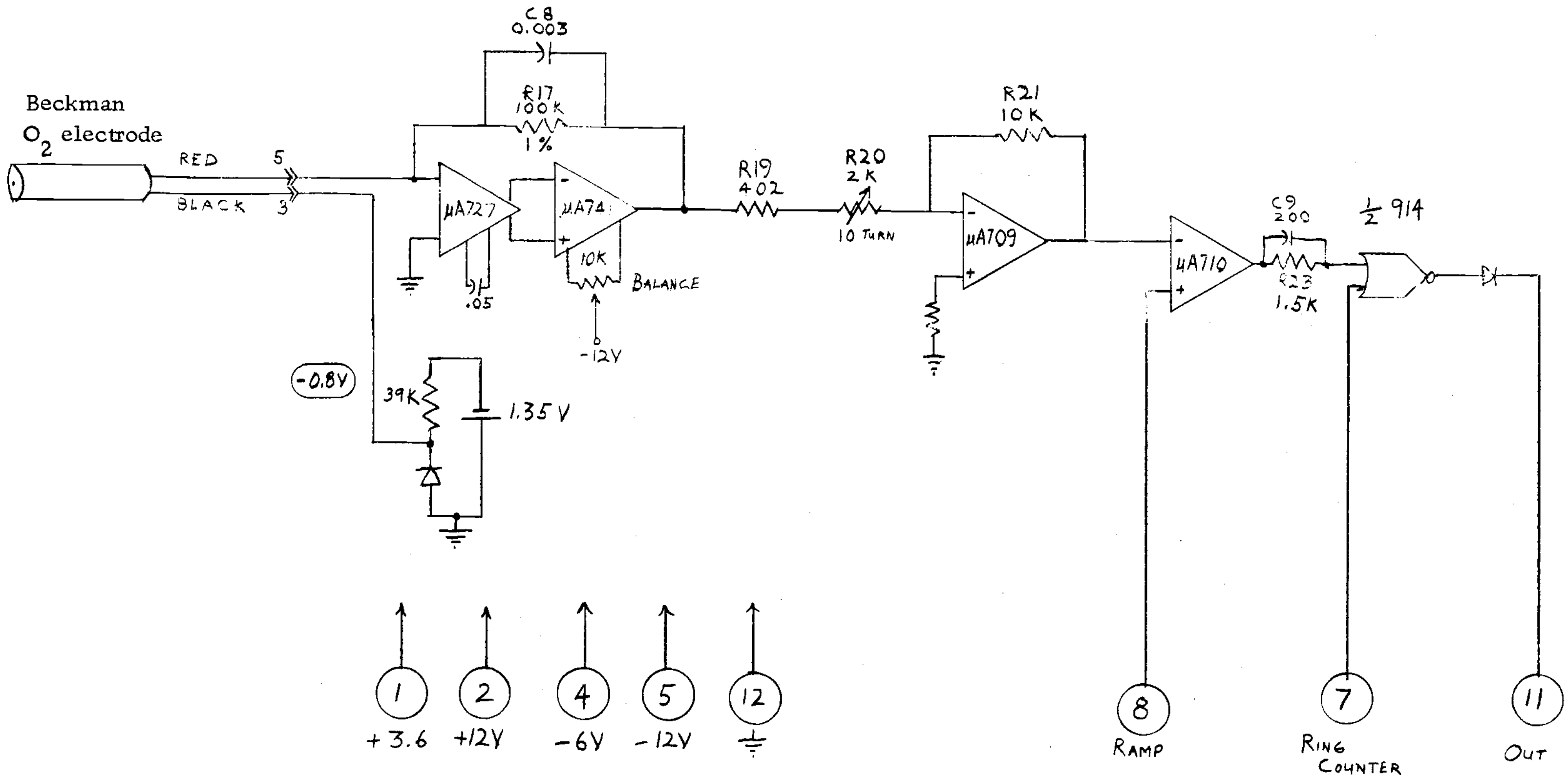


Figure 19. Signal conditioner for oxygen electrode.

The Oxygen Signal Conditioner

The multiplexing device requires an input of from zero to two volts, as this is the range of the ramp voltage. The output of the oxygen electrode is a current, varying approximately from zero to one microamp. Current is directly proportional to oxygen partial pressure. The oxygen signal conditioner, which acts as the interface between the electrode and the multiplexing circuitry, must produce a voltage within the proper range proportional to the current from the electrode.

The circuit shown on the previous page is a current to voltage transducer. The first stage of amplification is achieved with the μA727 - μA741 pair. The sum of currents into the amplifier input must be zero, so the current from the electrode must be exactly matched by the current through the feedback resistor. A voltage develops at the output of the μA741 which forces the proper current through the feedback resistor. The 727 is an ultra-stable preamplifier used because of the low level of the input current and the inherently high gain of the amplifier configuration. This first amplifier is followed by a voltage amplifier with a gain adjustable from -4 to -20 with ten turn potentiometer R20. The output of this amplifier is the necessary voltage level for comparison with the ramp.

The pulse generating circuit, because it is so simple, is included on the same circuit board as the oxygen signal conditioner. The amplifier output is fed to one input of a μA710 voltage comparator, and the other input of the comparator is connected to the ramp voltage. The comparator output is a logical 0 at the start of the ramp and switches to a 1 when the ramp voltage exceeds the amplifier output. The comparator output is NORed with the signal from one channel of the ring counter. As long as the ring counter is a 1 the output of the gate is 0. When the ring counter goes to 0 the gate output is 1 as long as the comparator output is 0. The result is a logical 1 whose duration in time is proportional to the current from the oxygen electrode and which appears if and only if the ring counter has a 0 on the proper channel.

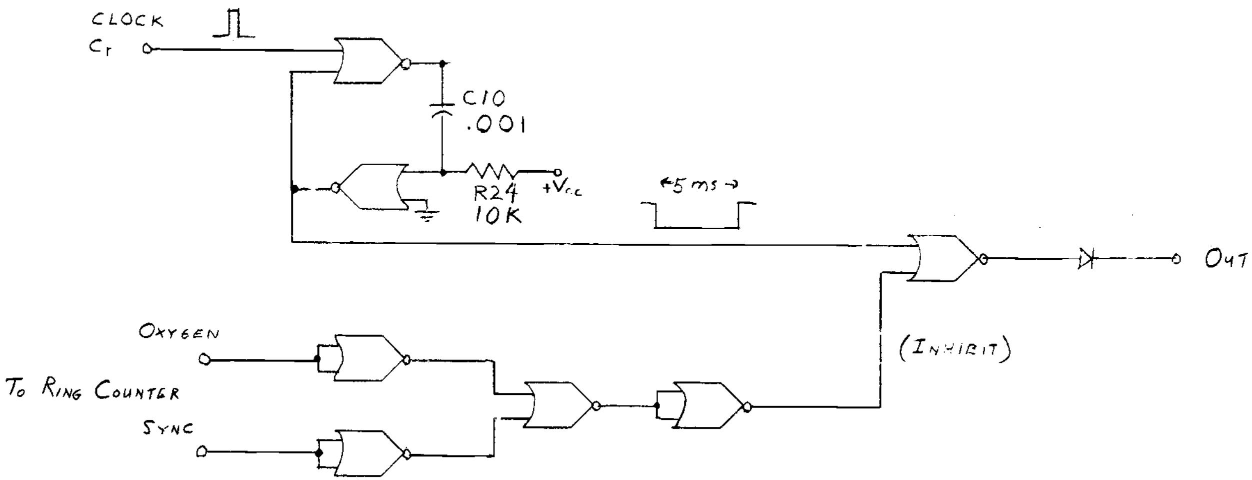


Figure 20. Channel marker.

Channel Marker

In the model constructed only the oxygen and synchronization channels are used. In any event, it is unlikely that all information channels will be used. Because the demultiplexer depends on receiving all pulses in order, it is necessary to inject a pulse into the unused channels. The circuit of Figure 20 does this.

The circuit consists of a pulse generator triggered by the system clock and some gates to inhibit the pulse during information-bearing channels. Every time the clock pulse occurs, a pulse a few milliseconds long is produced by the monostable. This pulse is fed directly onto the transmission line through a gate and a diode. The ring counter channels used to turn on the oxygen pulse and the synchronization information are combined and used to inhibit the pulse from the monostable during those channels.

Demultiplexing

At the receiving station, the train of transmitted pulses must be separated so that the pulse corresponding to oxygen level always appears in one channel, and that for any other variable in its particular channel.

A block diagram of the demultiplexer is shown in Figure 21. The pulse train from the receiver is fed to a Fairchild TTL 9301 decoder. The decoder also receives a three bit address from a counter. There are eight outputs on the decoder, and whichever output is addressed by the counter will have the pulses on it. The end of each pulse is used to increment the address by one, so that when a given pulse ends, the data is applied to the next output of the decoder. The data is applied to each of the eight outputs, one pulse on each, until the eighth output is reached. All further pulses will appear on the eighth output until a reset signal is received on the data line. This will cause the address counter to reset to 000, the address of the first output. Since there are eight pulses altogether, including the synchronizing signal, there should always be just one pulse per channel and the synchronizing signal should always appear on channel eight, resetting the device to channel one. The holding feature, which prevents a return to channel one without the proper synchronizing signal, is just a safety precaution in case an extra

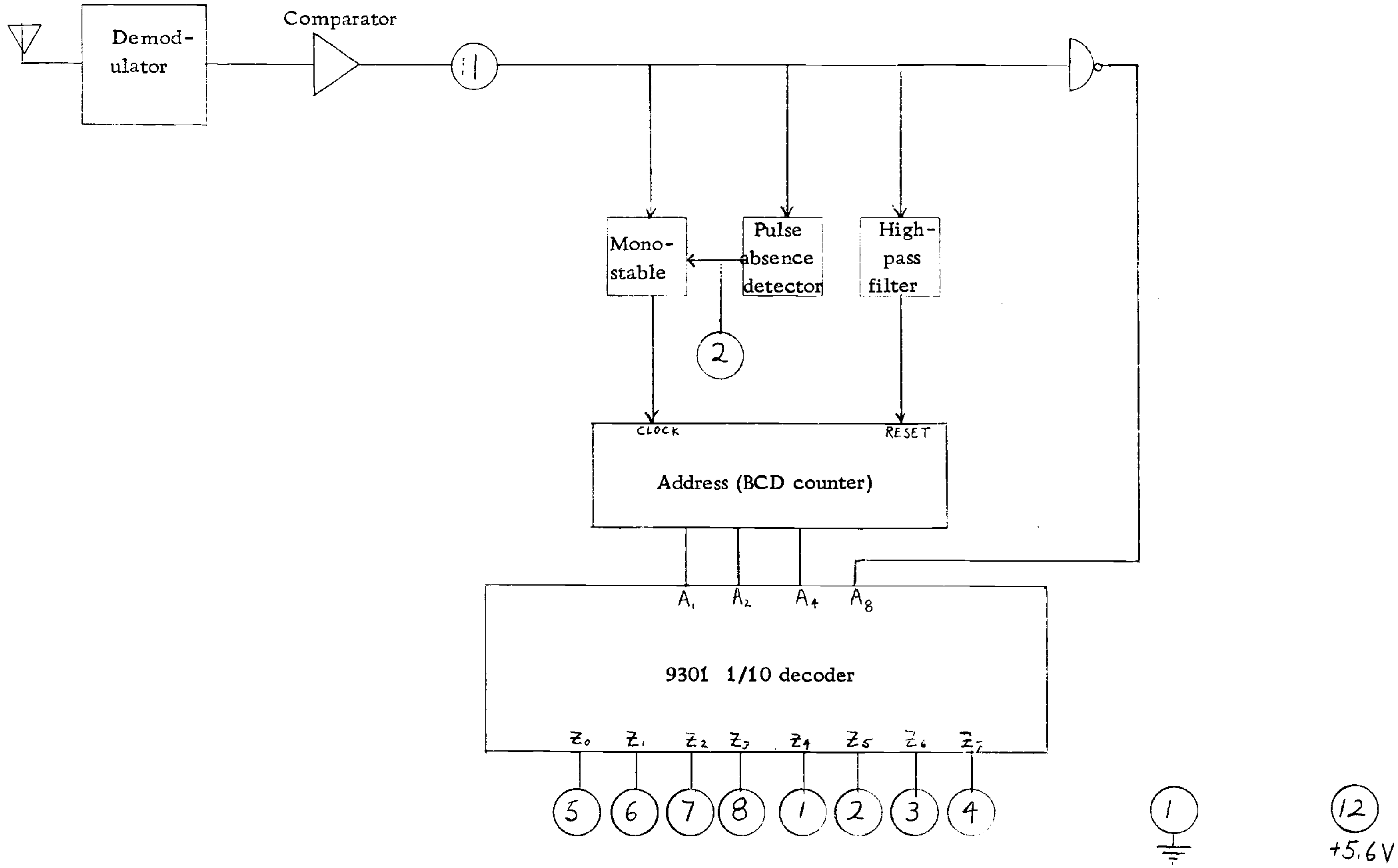


Figure 21. Demultiplexing, general scheme.

pulse is introduced into the data line by error.

The circuits on the following pages show the details of the demultiplexing circuitry. The counter clock pulse is generated by a monostable multivibrator, shown in Figure 22. The data line from the receiver is fed directly into the multivibrator input. Each time the data makes a 1 to 0 transition (at the end of each pulse), a short negative spike is produced at the multivibrator output. The resistor and capacitor on the multivibrator set the width of the pulse at one microsecond. This pulse is used to clock the counter.

The device labeled "9601" is a retriggerable monostable multivibrator. Its output is also connected to the clock multivibrator so that if the 9601 output falls from 1 to 0, a clock pulse is produced. The 9601 is ordinarily always in a high or triggered state. This is because it is triggered by the data line and just before it is ready to return to the "normal" 0 output, it is retriggered by a new pulse on the data line. This serves two purposes. If for some reason a pulse is missing in the pulse train the 9601 will provide a pulse in its place. The 9601 will also produce an output which can be used to light an indicator lamp in the event a pulse is missing from the data train. The pulses on the pulse train are ordinarily 12 milliseconds apart. The 9601 is set to provide a 15 millisecond output pulse. As long as a pulse appears on the data line every 12 milliseconds, the 9601 can not return to its normal state. If a pulse is missing, the 9601 will not be retriggered and will lapse into its untriggered state.

The address generator, illustrated in Figure 23, consists of three JK flip-flops. They are all clocked by the pulse from the multivibrator. The J and K inputs to the flip-flops are wired in such a manner as to cause them to count in ordinary binary fashion each time a clock pulse is received. When the state 111 is reached, the flip-flops are disabled so that no further change of state can occur.

A reset pulse returns all the flip-flops to 0's. The reset pulse is generated by a high pass filter followed by an integrator, as in Figure 24. The high pass filter consists of C12 and R28. The output of the filter is rectified and integrated on C13. The circuit of the diode, R29, R30, and C13 is actually a diode pump. Every positive pulse from the high pass filter causes a small amount of charge to go onto capacitor C13. This charge leaks off through resistor R30. If a large number of pulses appear in a short amount of time, the charge on C13 builds up faster than it can leak off. When the voltage across C13 reaches a sufficient magnitude, the transistor turns on, turning on the gate. The gate output falls to 0 until the charge leaks off the capacitor. This 0 level resets the flip-flops in the counter. The input pulses to the high pass filter must be at least 10 kHz to produce the reset pulse.

The 9301 one-of-ten decoder acts as a demultiplexer in the configuration used in this circuit (8). Binary numbers up to ten

(1010) can be addressed on the four input leads, causing the corresponding output lead to become a 0. All other outputs are 1's. If an address is placed on the first three address inputs and data is placed on the most significant address input, the device acts as a demultiplexer. If the data input is 0, the channel addressed by the other three inputs will become a 0. If the data input goes high, this is equivalent to addressing another output (above eight), so all outputs must be 1's. Because the unaddressed inputs are high, the pulse information must be low. Up to this point, the significant parameter has been the "on" or high duration of the pulse. For input to the decoder, the data line must be inverted. The significant quantity is then the low time of each of the several outputs.

As the last step in data processing, the low time of each channel must be translated into a voltage analog. The circuit which accomplishes this function, shown in Figure 25, consists of an integrator and a sample and hold circuit.

The integrator integrates a constant current on to C14 during the low time of the data input (one channel of the decoder output). The result of this operation is a voltage output which is a linear ramp during the pulse and levels off to a constant value when the pulse turns off.

The sample and hold circuit is activated during the channel following the data-bearing channel, to insure that the integration has been completed. The pulse from the following channel is used to turn on a field-effect transistor which allows capacitor C15 to charge to the same voltage level as the integrator output. This voltage level is sensed by an operational amplifier in a voltage follower configuration. The sampling circuit is enabled only during the pulse after the data-bearing pulse, so the voltage on the sampling capacitor C15 can change only during this time period. In the channel just preceeding the data-bearing channel, the integrator output is reset to zero through field effect transistor Q111. The resulting output of the voltage follower is a voltage level proportional to the on time of the data pulse, which is readjusted after each pulse is received and does not change in between pulses.

A low pass filter was originally tried on the pulsating output from the decoder, but in order to smooth out the approximately 10 Hz frequency of the output a very large time constant was required. The circuit worked, but over two minutes was required for the filter output to reach a steady state following a step change in the pulse width. Consequently, the filtering method was discarded in favor of the sample and hold system.

Experimental

The circuit boards for the transmitting portion of the system were placed in sockets in an aluminum box. The output bus, which

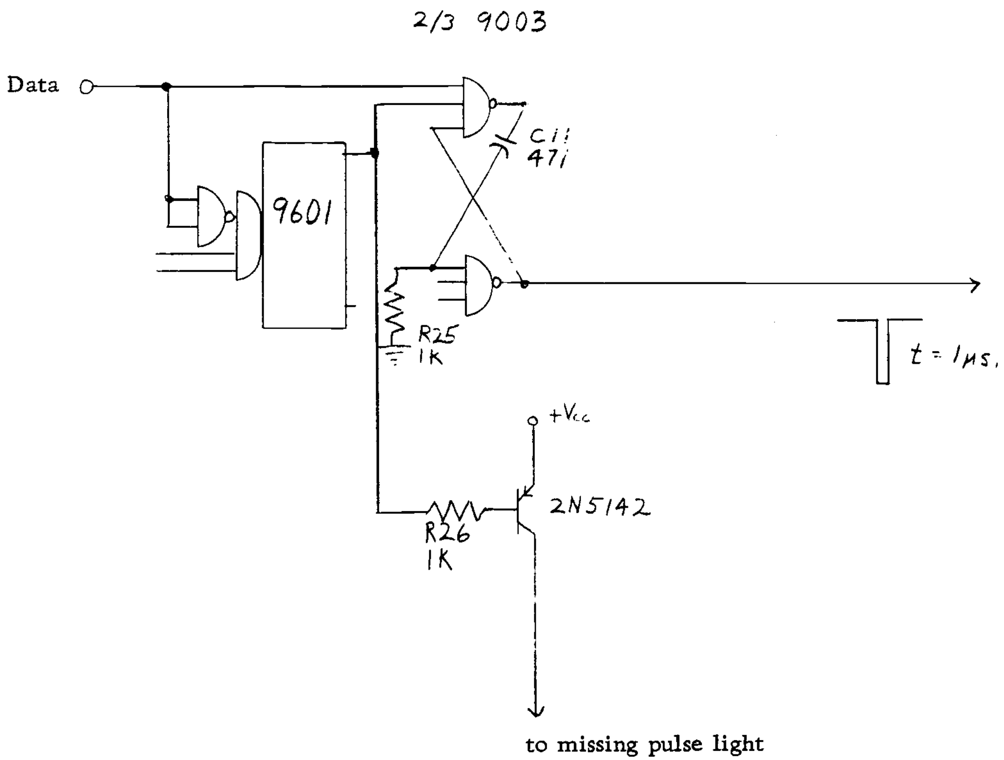


Figure 22. Clock for address generator.

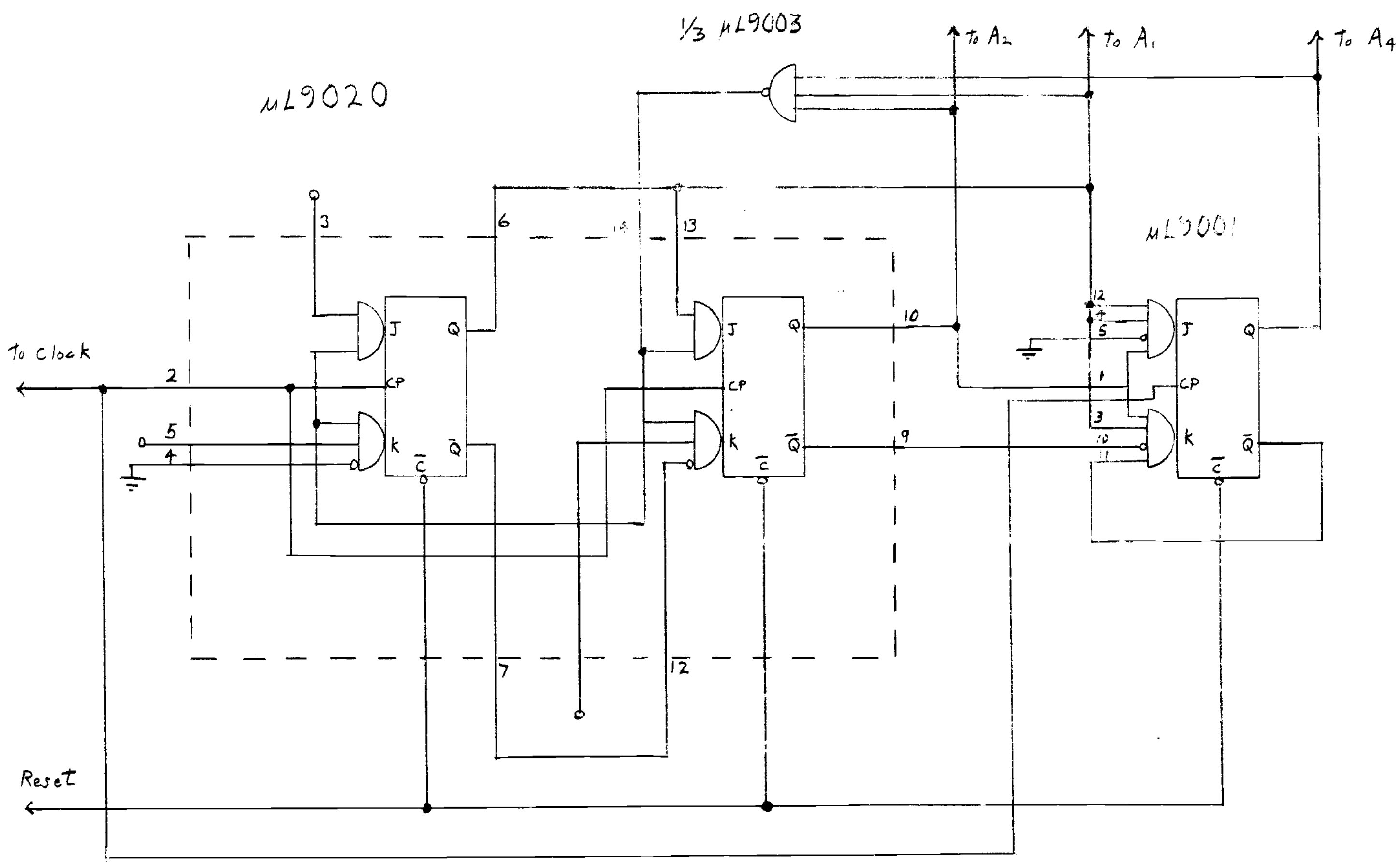
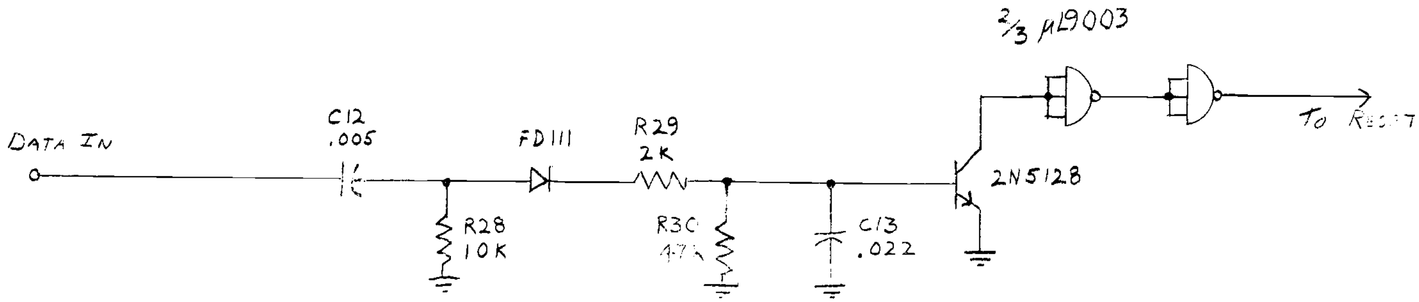


Figure 23. Address generator - up BCD.



Produces pulse on receiving 20 kHz signal.

[f(in) $\geq$ 10 kHz]

Figure 24. High pass filter and integrator.

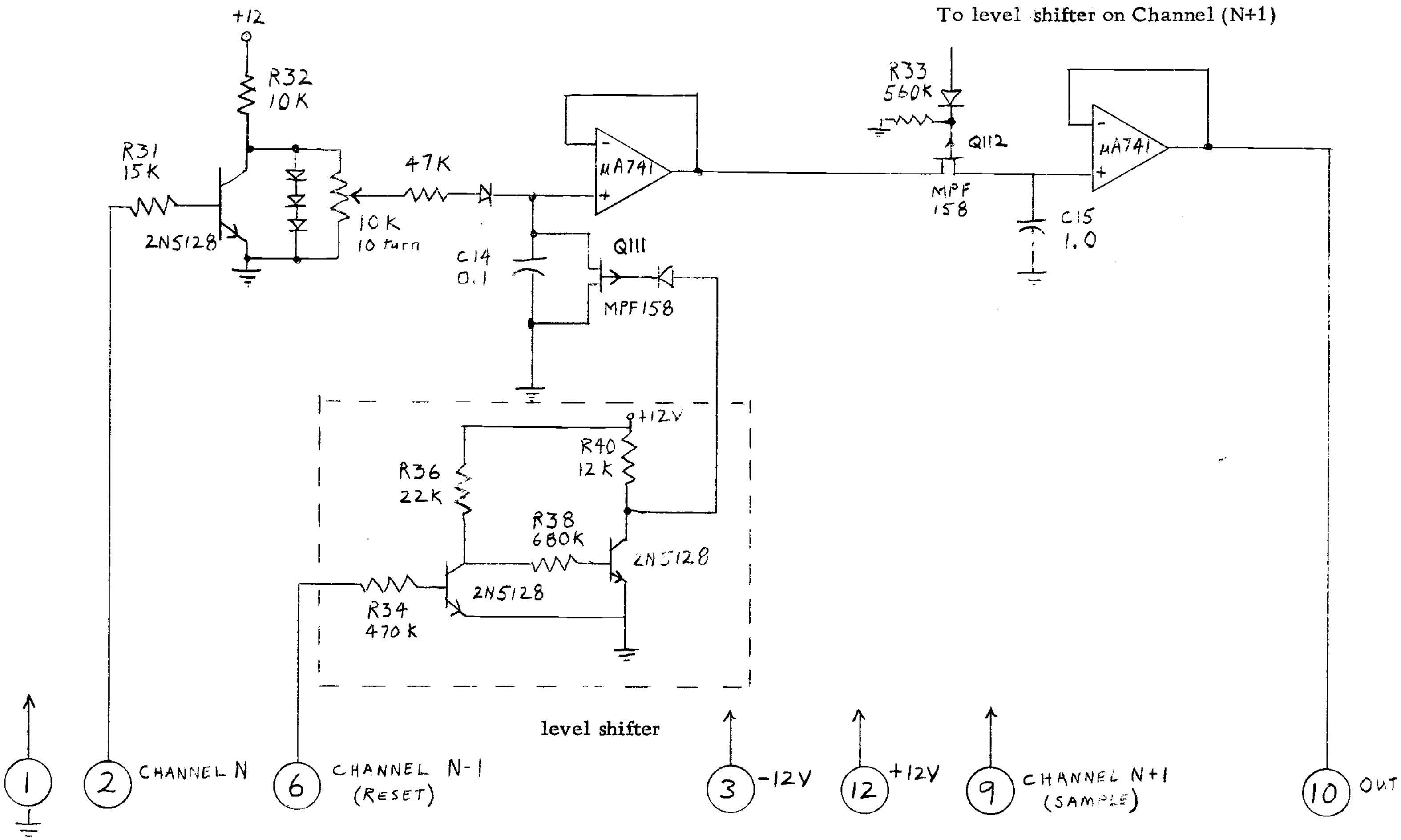


Figure 25. Integrate, sample, and hold.

in a radio link would be fed to the transmitter, was connected to a comparator. The other comparator input was biased at +0.6 volts. The comparator output was fed to the demultiplexing circuitry. The voltage from the sample and hold circuit was filtered (RC = 0.1) and read on a digital voltmeter.

Initially, the oxygen electrode was replaced with a current source. Figure 26 shows a graph of the output voltage of the current amplifier and also the output of the sample and hold circuit versus input current. The linearity of the current amplifier is excellent, but the sample and hold circuit shows non-linear response.

The oxygen electrode was replaced and readings were taken on several gases. Tanks containing 0%, 5%, 10%, and 15% oxygen in nitrogen were obtained from Salem Steel. These gases were bubbled through water to saturate them with water vapor and run past the oxygen electrode. The resulting calibration curve is shown in Figure 27.

Readings were taken on compressed air on a number of different days at different ambient temperatures. As expected, there was considerable temperature effect on the system, because the oxygen electrode is not compensated. The coefficient near 25°C is about 3.5%/°C. At constant temperature there was no detectable drift over a period of five hours.

The Avionics telemetry system used by the Department of

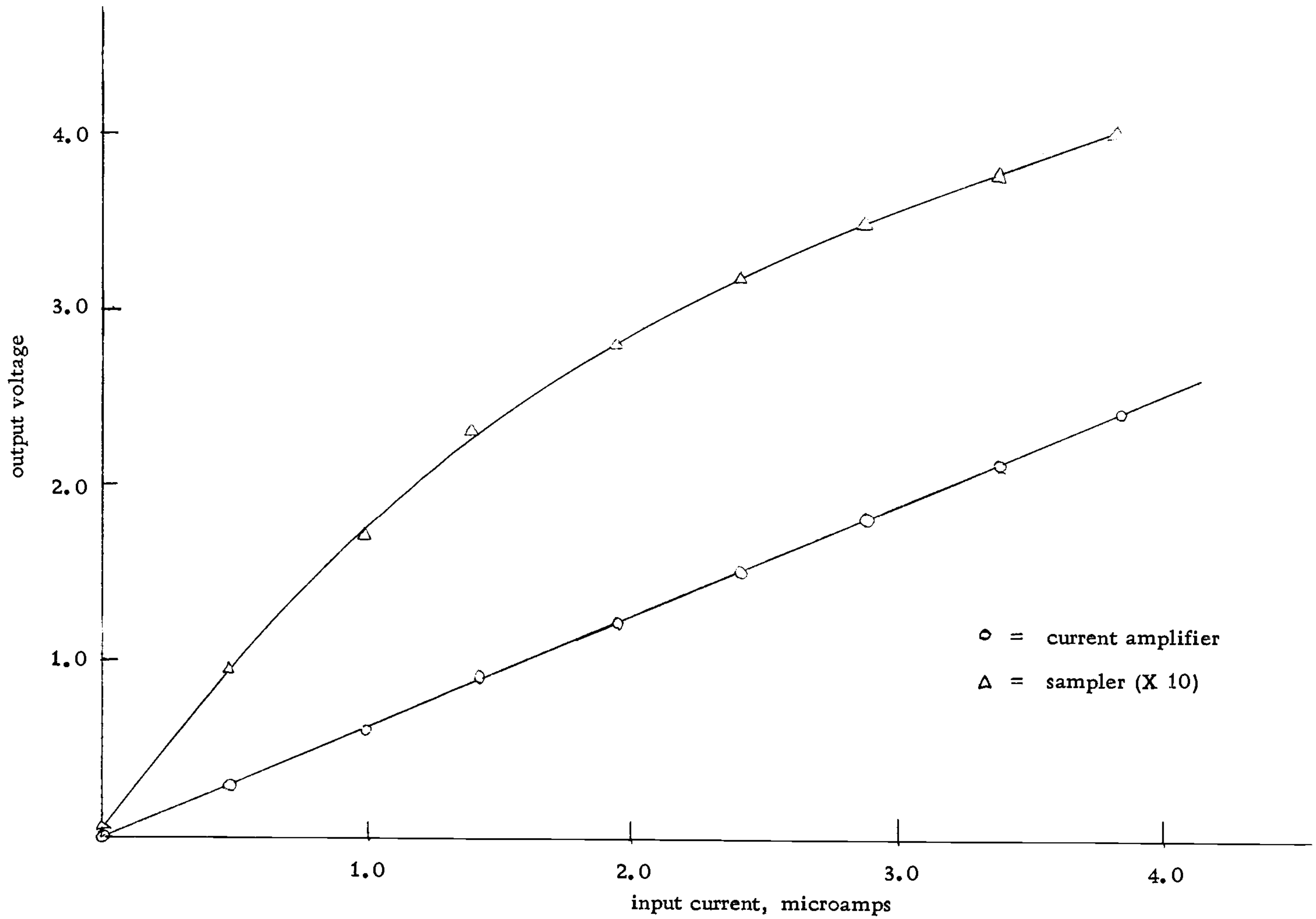


Figure 26. Response of system to input current.

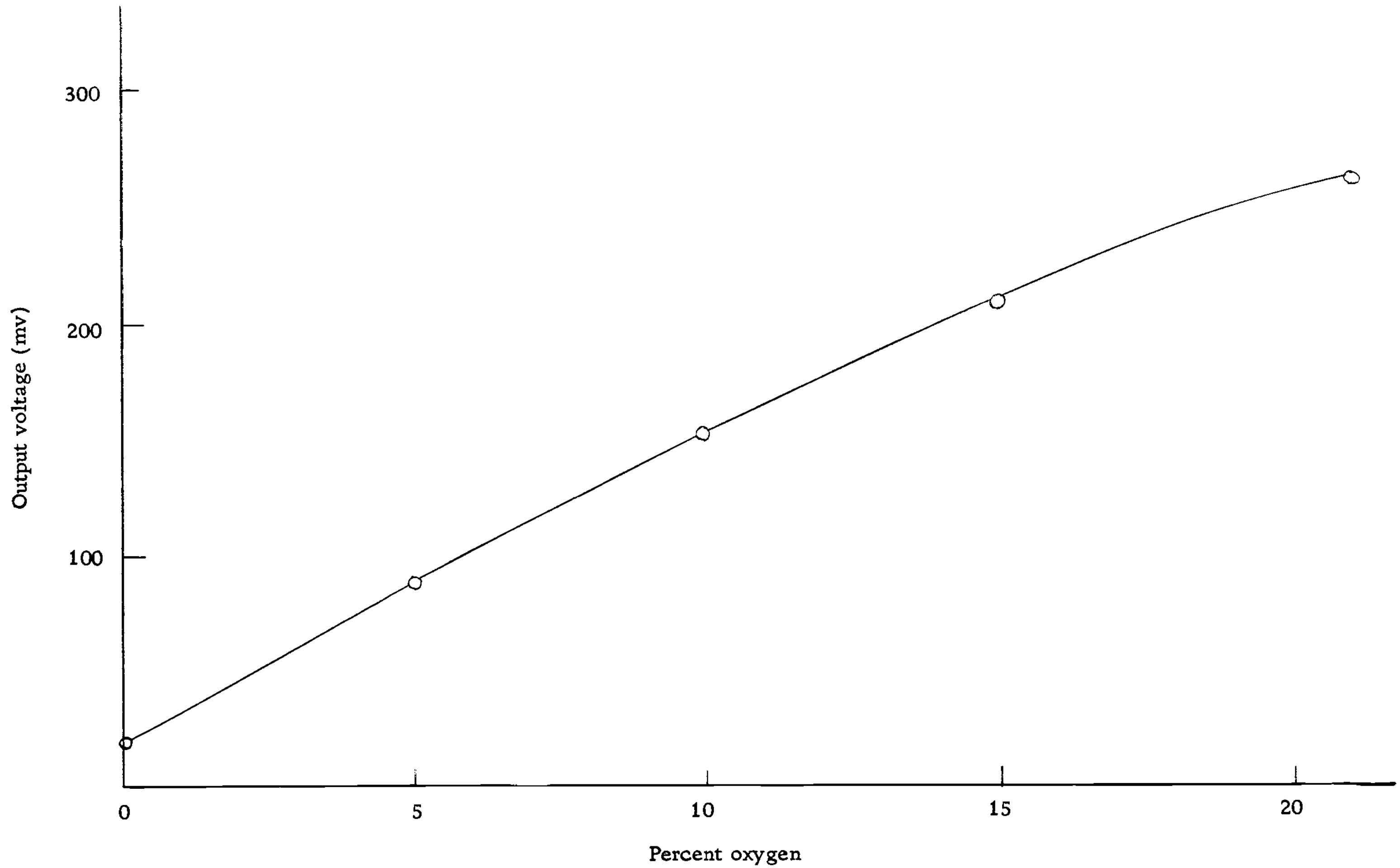


Figure 27. Calibration curve.
Ambient temperature, 30° C.

Physical Education at Oregon State was used in a radio link. However, the simple bio-medical telemetry transmitters which are available there are not suitable for transmission of the pulse information. The signal to noise ratio in the received signal is too small to make recovery of the pulse train possible. It will be necessary to construct an FM transmitter which can properly respond to the pulse train before the system can actually be applied to telemetry.

Some inconvenience is caused by the non-linearity in the sample and hold circuit. The non-linearity can be attributed chiefly to two factors. First, the capacitors used for integration and storage are not of sufficiently high quality. They should be replaced with computer grade polycarbonate types. The present devices exhibit fairly severe leakage. Second, both operational amplifiers need small bias currents at their inputs. If the bias current is not supplied externally, then the current is drawn from the capacitor. A current of about 10 nanoamps supplied to the first operational amplifier resulted in a marked improvement in performance. This bias current needs to be more closely optimized, however.

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APPENDIX I

LOGIC OPERATIONS AND SYMBOLOGY

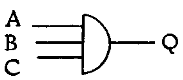
By making use of monolithic logic elements, one can design quite complex digital circuits without a real understanding of the switching characteristics of solid state devices. It is necessary only to have a knowledge of logical operations; integrated circuits can then be purchased which will implement the required logic. It is a good practice to know in advance the kinds of logic functions available in off-the-shelf integrated circuits. If digital systems are designed around these devices, great savings in cost and time can be realized over any other method of logic design.

Logic can be defined as either positive or negative. Positive logic means that the active or true level is high; negative logic means the true level is low. With standard RTL (resistor-transistor logic) elements, the logical high is about two volts; the logical low is less than half a volt. Using the positive logic convention, the two volt level is defined as 1 and the half volt level is defined as 0. Logic circuits are described according to how they respond to logical 1's at their inputs.

The basic functions in logic operations are AND and OR. The output of a two input AND gate is high (1) if and only if both of its inputs are high. The output of an OR gate is high if and only if either one or both of its inputs are high.

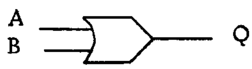
The response of a device to its inputs is frequently shown in a truth table. Illustrated below are a three input AND gate and a two input OR gate, with their truth tables. Notice the difference in the logical symbols.

AND



A	B	C	Q
0	0	0	0
0	0	1	0
0	1	0	0
0	1	1	0
1	0	0	0
1	0	1	0
1	1	0	0
1	1	1	1

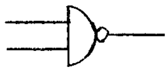
OR



A	B	Q
0	0	0
0	1	1
1	0	1
1	1	1

In actual practice, it is found that the inverses of these functions are more practical. The inverse functions are called NAND and NOR and merely indicate that the output has been inverted. A NAND gate produces a logical 0 if and only if all of its inputs are 1's. A NOR gate produces a logical 0 if any of its inputs are 1's. The inversion is shown on the symbol by a small circle at the inverted lead:

NAND



NOR

Writing of logical equations and manipulation of logic operations has been developed from Boolean algebra. The currently accepted notation for OR is a plus sign. AND is indicated by a dot as in multiplication or by writing symbols in conjunction with no operation symbol at all. Hence the AND and OR gates illustrated above can be described by the following two equations:

AND

$$Q = A \cdot B \cdot C$$

or

$$Q = ABC$$

OR

$$Q = A + B$$

Inversion is indicated in equations by a bar over the variable to be inverted. The NAND equation is

$$\overline{Q} = ABC \qquad \text{or} \qquad Q = \overline{ABC}$$

The second form of the equation is evident if the truth table for the gate is examined. The output is exactly opposite that for the OR gate. The second form of the equation can also be derived algebraicly by negating both sides of the first equation using the following rules:

If $X = Y$, then $\overline{X} = \overline{Y}$

$\overline{\overline{X}} = X$ X and Y are any logical expressions.

Inversion is sometimes symbolized by a prime mark. This notation has a slightly different meaning than the bar. $\overline{X}$ generally means $X = 0$; X' means that the variable so marked is opposite to X. The result of either notation is the same; only the connotation is different.

One form of the equation for the NAND gate was given as $\overline{ABC}$. This is not the same as $\overline{A} \cdot \overline{B} \cdot \overline{C}$. The former expression means "not (A and B and C)", the latter means "not A and not B and not C." A Boolean Theorem known as DeMorgan's Law can be applied to convert an expression with a bar over the entire expression to one with bars over the individual elements of the expression.

$$\overline{ABC} = \overline{A} + \overline{B} + \overline{C}$$

also, $\overline{A + B} = \overline{A} \cdot \overline{B}$ A, B, and C, are any logical expressions.

Virtually any logical function can be generated by the appropriate use of NAND and NOR gates, along with inversion occasionally. One other function, which can be generated with the NAND and NOR, but which is itself sometimes considered a basic expression, is the EXCLUSIVE OR. The logical OR includes AND. That is, "A or B" means either A or B or both. The EXCLUSIVE OR (or "Cluse") excludes the AND portion. The symbol is a plus sign with a circle around it.

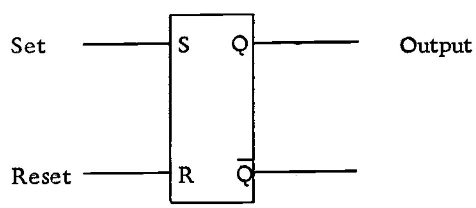
A	B	C
0	0	0
0	1	1
1	0	1
1	1	0

$$C = A \oplus B$$

The CLUSE function is important in computational circuits because it is the linear digital operation of addition. An EXCLUSIVE OR gate is sometimes called a half-adder. A full adder includes a third input and a separate AND output to accept and generate carry signals in addition operations.

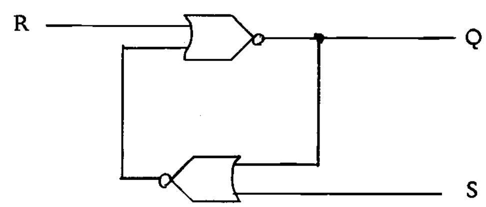
Another basic kind of logic circuit is the flip-flop. As the name implies, a flip-flop is a circuit with two discrete output states and it flips from one to the other on receiving a particular input signal.

The simplest kind of flip-flop is a SET-RESET, or R-S type. The circuit consists of two inputs, labeled set and reset, and an output (or two complementary outputs).



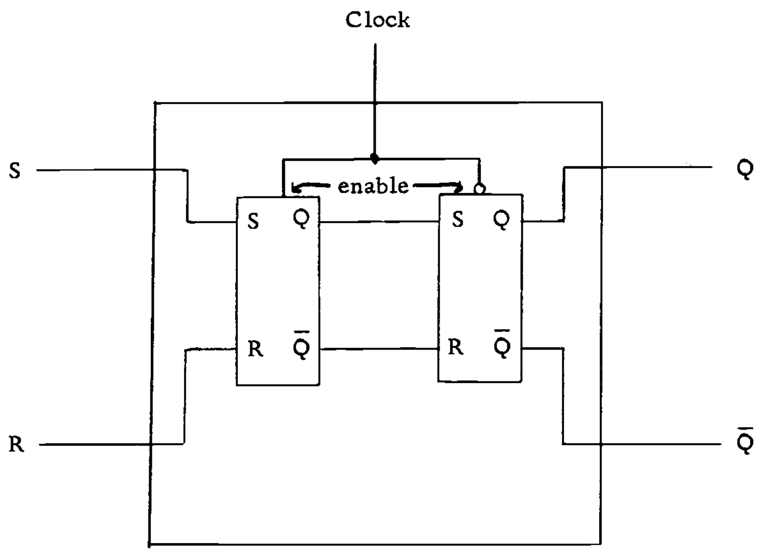
A 1 on the set input causes a 1 to appear on the output. A 1 on the reset input causes a 0 to appear at the output. Once the device has been set or reset, it will remain in the appropriate state until a signal causing it to change state is received. The flip-flop then is essentially a memory element; it remembers what the last signal it received was.

The R-S flip-flop can be simply implemented with two NOR gates. Examination will show that a 1 on the lead labeled "S" will cause the output to become a 1, and that this state is stable. An input on the "R" lead will cause the output to change to a 0, and this state is also stable.



A flip-flop purchased as a monolithic logic element is more complex. There is an additional lead for a clock input. The flip-flop can change state only during some special state on the clock input. This allows the designer to set up some function at the inputs to the flip-flop through logic nets without affecting the output until the specified clock function occurs. For some flip-flops, the output changes only when the clock input changes rapidly from a logical 1 to a 0. The Fairchild μL923 is such a device. The clock lead is capacitively coupled to the rest of the circuit, and a very fast fall time on the clock pulse is necessary to trigger it.

Some other flip-flops are "master-slave" configurations. The clock input is normally high. While the clock is high the inputs of the device are connected to the master flip-flop, which sets up according to the signals at its inputs. When the clock changes to low, the master flip-flop is disabled, so that it cannot change state, and its output is connected to a slave flip-flop, whose output changes to correspond to the output of the master flip-flop. When the clock goes back to high, the master is re-enabled, and the slave is disabled. These devices are generally operated with clock inputs consisting of periodic short high-to-low-to high transitions.



The drawing above is a symbolic representation of the master-slave flip-flop. The master is enabled by the high clock level, and the slave by the low clock level, as indicated by the inverting circle on the enable input.

There are other kinds of flip-flops beside the R-S type. A type D flip-flop has only one input, the data input, and whenever a clock pulse occurs, the output changes to correspond to whatever was on the D input.

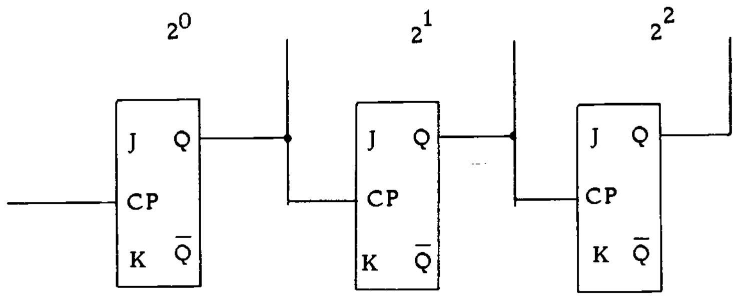
A type T flip-flop has only one input also. A 1 on the T, or trigger, input simply causes the device to change state. If the output was a 1, a input on T will cause it to change to a 0, and conversely.

A type similar to the R-S flip-flop is the J-K flip-flop. The R-S flip-flop has an undefined state possible, namely R = S = 1. This situation is not allowed on an R-S device. A J-K flip-flop behaves exactly like an R-S, except that the undefined state is allowed, and is defined to mean "toggle." That is, a J-K acts just like an R-S except that if J = K = 1 it acts like a type T. This operation is extremely useful, and most integrated circuit flip-flops are of the J-K type. The operation of the J-K flip-flop is summarized in the table below. The superscripts n and n+1 refer to times before and after a clock pulse, respectively.

$(J)^n$	$(K)^n$	$(Q)^n$	$(Q)^{n+1}$
0	0	0	0
0	0	1	1
0	1	0	0
0	1	1	0
1	0	0	1
1	0	1	1
1	1	0	1
1	1	1	0

$$(Q)^{n+1} = (\bar{K}Q + J\bar{Q})^n$$

Flip-flops can be used in two different ways to make counters which count up in BCD. If μL923 flip-flops are used, the output of each flip-flop is connected to the clock input of the next flip-flop.



This circuit makes use of the fact that for the uL923, a 0 on both the J and K inputs causes the device to toggle (rather than a 1, as is the norm). Each time a flip-flop output makes a transition from 1 to 0, the following flip-flop will toggle. Such a device is called a ripple counter, because each stage changes state only after the previous stage has changed.

Sometimes the delay in setting up the count due to the sequential nature of the ripple counter cannot be tolerated. In such a case a synchronous counter is used. In this circuit, all the flip-flops are clocked together and the correct state is set up by setting proper signals at the J and K inputs of the flip-flops. The algorithm for the inputs is based on the fact that any time a flip-flop is to change state in a BCD code then all the less significant bits must be 1's. To make a BCD counter then, each flip-flop has its J and K inputs tied together and connected to an AND gate over all the previous bits of the counter. Then each flip-flop will toggle when all previous bits are 1.

In the equipment constructed for this research, the BCD counter used to address the decoder in the demultiplexer is a synchronous counter.

APPENDIX II

LINEAR INTEGRATED CIRCUITS

The availability of monolithic operational amplifiers for only a few dollars per unit makes possible the construction of fairly complex precision instruments at low to moderate costs. There are some special characteristics of integrated circuit amplifiers which deserve mention.

As a rule, a transistor circuit has a lower input impedance than a tube version of the same circuit. Impedances in monolithic circuits are lowered further because of the difficulty of producing large resistances on the chip. In operational amplifier circuits, the input impedance of the amplifier is important because most of the transfer functions assume a negligible current flow into the amplifier. As long as the amplifier input impedance is much greater than the other impedances in the circuit (feedback resistor, input source resistance), then the negligible input current assumption is satisfied. Tube type operational amplifiers may have input impedances on the order of tens of millions of ohms or more. Discrete transistor operational amplifiers typically have input impedances of around a megohm. Integrated circuit units such as the 709 type have impedances of about 300,000 ohms.

Consequently, it is usually necessary to consider the non-ideality of the amplifier when integrated circuits are used. Impedances can be raised by use of a voltage follower configuration or by use of field effect transistors in the input. Several manufacturers are now marketing or plan to market in the near future special amplifiers for instrumentation with built in field effect transistors or with other special circuitry to greatly improve input impedance. Notably, these are Analog Devices, Inc., National Semiconductor, and Fairchild.

Another difficulty encountered with monolithic operational amplifiers is directly associated with one of their great advantages. The gain of the amplifier is very high and constant over a large frequency range. This characteristic creates stability problems. The so-called first generation amplifiers require external components to limit the gain at high frequencies. The 709 type amplifier has two points at which frequency compensation must be applied. The compensation is accomplished by adding resistors and capacitors whose values are determined by the desired closed loop gain of the circuit. If these components are not used, or are improperly selected, the amplifier may be expected to oscillate at high frequency.

The second generation operational amplifiers, such as Fairchild's uA741, has the frequency compensation built in, and does not require any external compensation. The uA741 is a direct plug-in replacement for the 709.

Even with frequency compensation, the amplifier may not be stable. Oscillations can occur due to coupling between amplifiers through the power supply leads. Every amplifier should have its power supply leads bypassed to ground through a 0.01 μF disc capacitor at the amplifier terminals.

If the above mentioned precautions are observed, the monolithic operational amplifier is an extremely valuable tool. Its greatest asset is probably cost, but there are other features worth pointing out. The gain of the amplifier is typically 100 dB, flat to at least 10 kHz. This means that the amplifier can handle input signals from D.C. to moderately high frequency with no change in gain. The noise levels at the inputs and the drift in input bias currents vary widely from one model to another, but as a rule are remarkably low. The insensitivity of the inputs to common-mode voltages and the stability of the gain over wide temperature ranges cannot be equalled by another kind of amplifier except at expense of hundreds of dollars. Lastly, for portable equipment, the low power consumption and small size of the monolithic amplifier make possible instrumentation which otherwise could simply not be done.

There are several references mandatory for anyone contemplating use of monolithic operational amplifiers. These are the

Philbrick manual (7), the Fairchild applications manual (9), and the collection of papers edited by Eimbinder (4).