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R. L. E. TECHNICAL REPORT 413

A MULTICHANNEL ELECTROENCEPHALOGRAPHIC TELEMETERING SYSTEM

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TECHNICAL REPORT 413

NOVEMBER 6, 1963

MASSACHUSETTS INSTITUTE OF TECHNOLOGY
RESEARCH LABORATORY OF ELECTRONICS
CAMBRIDGE, MASSACHUSETTS

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The research reported in this document was made possible in part by support extended the Massachusetts Institute of Technology, Research Laboratory of Electronics, jointly by the U. S. Army (Signal Corps), the U. S. Navy (Office of Naval Research), and the U. S. Air Force (Office of Scientific Research) under Contract DA36-039-sc-78108, Department of the Army Task 3-99-25-001-08; and in part by Grant DA-SIG-36-039-61-G14; additional support was received from the National Science Foundation (Grant G-16526) and the National Institutes of Health (Grant MH-04737-03).

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A MULTICHANNEL ELECTROENCEPHALOGRAPHIC TELEMETERING SYSTEM

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This report is based on a thesis submitted to the Department of Electrical Engineering, M.I.T., May 17, 1963, in partial fulfillment of the requirements for the degree of Master of Science.

(Revised manuscript received August 26, 1963)

Abstract

A four-channel electroencephalographic telemetering system small enough to be carried by an animal of the size of a cat has been designed and built.

Descriptions of the differential amplifiers, pulse-duration multiplexing circuitry, transmitter and receiver, and decoding circuitry are given here. Details for reproduction of the system, and a simple module method for adding more channels are described in the appendices.

Each channel of the four-channel system has a frequency range 0.7-2000 cps with a noise level referred to the input of 5 μv peak-to-peak at 10,000 ohms. The over-all operating range of the system is 200 feet.

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I. INTRODUCTION

Electroencephalographic (EEG) signals from humans and animals are usually recorded from subjects at rest or immobilized. In certain experiments, such as behavioral studies, it is desired to monitor the EEG of unanesthetized subjects without these restraints. One technique is to use a long cable between the subject and the recording apparatus. Cables, however, often introduce more difficulties partially because the animal is still restrained to some extent. Cables also introduce artefacts arising mainly from the change of the electrostatic charge of the insulating material of the wire and induction of an emf in the wire loop as a result of its movement in the earth's magnetic field.[1]

If, however, the EEG is converted to radio waves and telemetered to the receiver and recording apparatus, these difficulties are overcome. The idea of using telemetry for EEG is not new, but only recently, with the advent of subminiature transistors with high gains and associated subminiature components, has it become truly practical.

Several portable EEG amplifying-transmitting devices using vacuum tubes have been developed.[2-6] These are all single-channel devices and suffer from weight and power requirements. Single-channel transistor telemetering systems have been developed,[1,7-10] and a two-channel system has been described by Kamp.[11] The latter system, although having two channels, cannot conveniently be extended beyond three-channel operation.

A modification of Kamp's system was built at M.I.T.[12] Although it telemeters four channels simultaneously, it suffers from the need of two radio transmitters, which often interact. Litton Systems, Inc., Los Angeles, Calif., has marketed a three-channel telemetry system. However, only one of these channels has a bandwidth that is capable of handling EEG signals.[13]

The need for multichannel operation with flexibility in adding more channels was expressed by Kamp. He has recently described an eight-channel EEG telemetering system.[14] Each channel of this system has a frequency response of 1-100 cps. However, there are certain applications, as pointed out in Section II, in which this frequency response is not sufficient. The high cost of receivers and the problems of transmitter coupling and interaction make the use of several single-channel systems, to obtain multichannel operation, impractical.

There is a need for a multichannel EEG telemetering system for studying cortical and subcortical evoked responses in unanesthetized, freely moving animals. The four-channel system described in this report was designed for experiments with unrestrained cats. This system is small and light enough to be strapped on the back of an animal of the size of a cat, requires only one radiofrequency (rf) transmitter, and can easily be extended to more channels, all of which are capable of telemetering cortical and subcortical evoked EEG.

Against these advantages of telemetry must be weighted certain disadvantages. The complexity and the cost of telemetry is greater than direct-wire recording. Also, artefacts that are due to rf noise and loss of rf signals as a result of the subject's position with respect to the receiving antennae are occasionally encountered. With respect to multichannel operation, it must be remembered that the weight and power requirements increase as the number of channels increases.

Although telemetry does eliminate artefacts that result from long cables, it does not decrease artefacts caused by transmission of muscle activity and shifting of electrode contacts. These can be reduced, however, by certain techniques of implanting electrodes.[1]

II. DESIGN SPECIFICATIONS AND THE BASIC SYSTEM

An attempt to design a telemetry system that would handle all of the possible electrical signals produced by the nervous system would be extremely difficult, if not impossible. Even if a certain area of the nervous system is to be studied with only gross electrodes, the specifications of the telemetry system become dependent on many factors, such as whether or not stimulation is to be used, the type of stimulation, the type of electrodes, and where the electrodes are placed.

For example, on-going activity (recorded in the absence of intentional stimulation), recorded from gross electrodes on the surface of the cortex, usually does not have frequency components above 200 cps. This upper frequency limit generally holds true even if "physiological" stimuli, such as optical or acoustical stimuli, are used. However, if the same gross electrodes are used, but shocks are delivered to subcortical structures, frequencies up to 2000 cps may be present. Still higher frequencies are encountered (up to 5000 cps or higher) if subcortical recording electrodes are used, especially if they are microelectrodes or semimicroelectrodes.

The amplitude of the signals present also covers a large range. In studies in which averaging techniques were used consistent signals of approximately 5 μv have been reported.[15] On the other hand, if microelectrodes are placed in single cells, the signals may be as great as 100 mv.

One anticipated use of this system is to study responses to sensory stimuli and shock stimulation of sensory pathways recorded from gross electrodes on the surface of the cortex of cats in different physiological states. Thus the specifications must cover this application. To allow the subject essentially unrestricted movement, an operating range of at least 100 feet is necessary, and yet battery drain should be kept low enough for at least 10 hours of uninterrupted service. The number of simultaneous channels should be at least four and should be easily extendible for future expansion of the system.

With these specifications in mind, the design criteria for the four-channel system were set as follows.

1) Each channel of the system should have a frequency response of ±3 db in the range 1-2000 cps. Although this falls short of the total frequency range that is possible, it does cover the specific application for which this system is designed and is more than adequate for conventional gross-electrode studies of on-going activity and evoked responses from "physiological" stimuli.

2) The system should telemeter four channels simultaneously on one radiofrequency.

3) The telemetering package should be immune to normal shock and should operate satisfactorily over 20-40°C.

4) Battery life should be at least 12 hours between rechargings.

5) The over-all noise level referred to the input must be less than the smallest signals encountered, that is, 5 μv.

More specific requirements, for example, input impedance of amplifiers, are

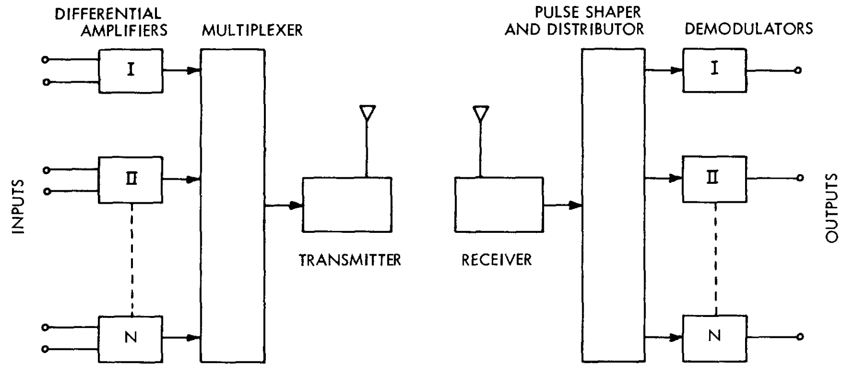

Fig. 1. Block diagram of EEG telemetry system.

discussed in other sections.

A block diagram of the system is shown in Fig. 1.

The low-noise, broadband, high-gain differential amplifiers provide signals of the proper level for the multiplexer. The function of the multiplexer is to combine the different signals from the amplifiers into a single signal that then modulates an FM transmitter in the range 88-108 Mc.

The received signal, after detection by an FM receiver, is divided into its original signals by the pulse shaper, the distributor, and the demodulators. The outputs can be connected directly to a pen recorder or an oscilloscope.

III. THE DIFFERENTIAL AMPLIFIERS

As previously mentioned, the EEG signals of interest have a possible dynamic range from 5 μv to 100 mv, that is, 86 db. Since it is difficult to design an amplifier with this dynamic range with a low supply voltage, not to mention the difficulties that the multiplexer would have in handling such a range, it was decided to design a basic amplifier that would handle input signals in the range 5-250 μv (34 db). If signals greater than 250 μv are present, an attenuator can be added in the input.

The source impedance associated with these signals is 5000-50,000 ohms. To avoid loading of the signal, the amplifier input impedance must be much higher than 50,000 ohms. The design criterion was set at 500,000 ohms or greater. The frequencies of interest are in the range 1-2000 cps. The lower frequency bound proved to be the difficult one to achieve with an ac coupled amplifier.

It is anticipated that the system will be operated in the presence of undesirable common-mode signals. There are several methods of obtaining common-mode rejection. The simplest is a center tapped transformer input to a conventional single-ended amplifier. This transformer, however, is bulky and has poor low-frequency response.

The use of a differential amplifier results in good common-mode rejection but requires almost twice the number of components as a single-ended amplifier. However, the differential amplifier can be designed for high differential-mode gain and, at the same time, provide excellent dc operating-point stability. An ac coupled differential amplifier was designed (Fig. 2) which was packaged in less than 0.75 cubic inch by using readily available subminiature components.

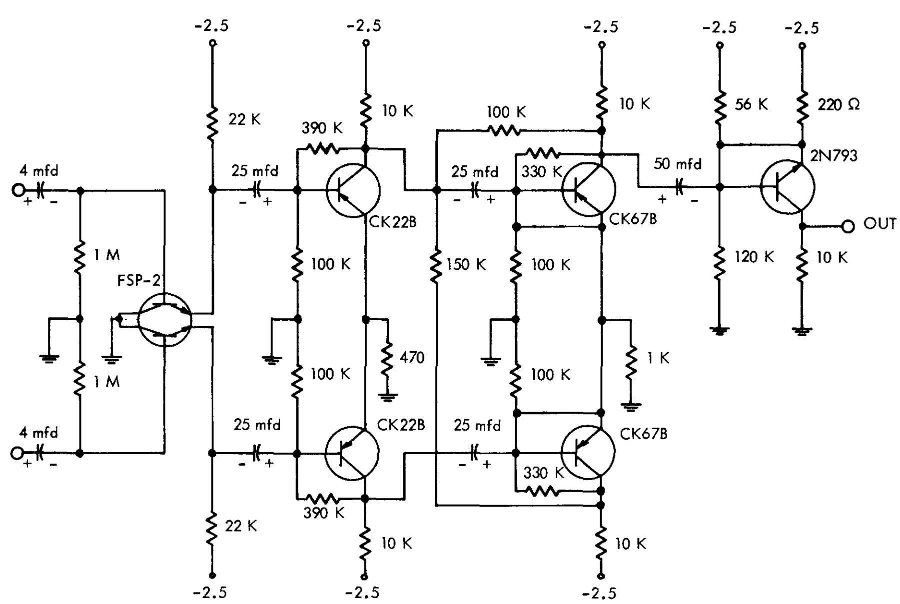

Fig. 2. Differential amplifier.

Table 1. Differential amplifier characteristics.

Voltage gain	4000 (differential input, single-ended output)
Dynamic range	34 db (5-250 μv peak-to-peak)
Frequency response	±3 db, 0.7-20,000 cps
Common-mode rejection	48 db at 60 cps 46 db at 500 cps 46 db at 2000 cps
Noise	Equivalent input noise referred to input (10-kohm source); 5 μv peak-to-peak over a bandwidth of 0.2-20 cps (1/f noise); 5 μv peak-to-peak over a bandwidth of 80-2000 cps
Input impedance	640 kohms
Output impedance	10 kohms
Power requirements	-2.5 volts at 0.6 ma

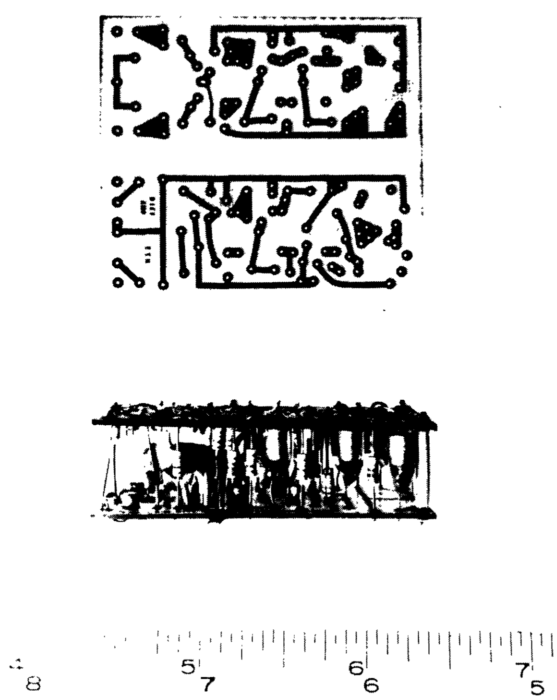

Fig. 3. Amplifier and associated multiplex module.

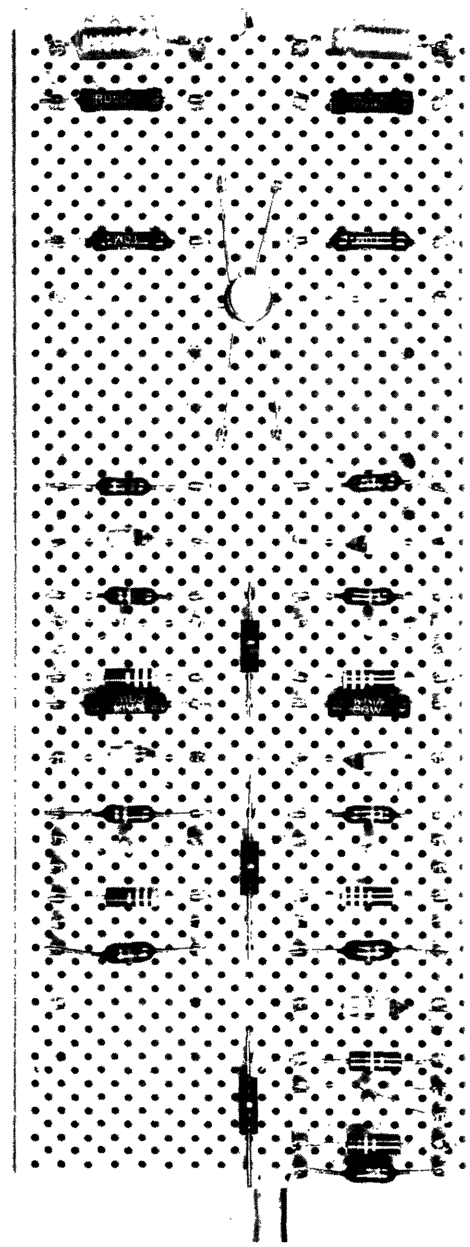

Fig. 4. Breadboard of differential amplifier.

The first stage is an emitter follower that is used as an impedance converter. The two silicon planar transistors are closely matched with respect to β, I_{ceo}, and $\Delta(V_{BE1}-V_{BE2})$ from 20-40°C and are packaged in the same can so that their temperatures will be approximately equal. The low leakage of silicon transistors allows the use of large emitter resistors which results in a high input impedance. The first stage is the source of most of the noise generated by the amplifier itself. This is kept low by using a low-noise FSP-2 dual planar transistor.

The second stage consists of a pair of CK22B transistors, which are low-noise subminiature transistors. These transistors are matched within 5 per cent with respect to β and I_{ceo}. Specific details concerning matching are contained in Appendix B. The capacitors coupling the first stage to the second are subminiature tantalums.

The third stage is similar to the second except that CK67B transistors are used. These are subminiature high-gain transistors. They are also matched within 5 per cent with respect to β and I_{ceo}. A form of negative common-mode voltage feedback that was suggested by Kamp is used in this stage, in addition to the conventional negative feedback from the collector to the base of each transistor. This feedback consists of voltage feedback from both collectors of the pair to the base of one of the transistors through the input coupling capacitors. The phase relationship between common-mode and differential-mode signals is such that the feedback greatly degenerates common-mode voltage gain but has little effect on differential-mode voltage gain. This additional feedback loop increased the common-mode rejection of the entire amplifier by 28 db, even with 10 per cent resistors in the feedback loop. Much higher common-mode rejection can be obtained if these resistors are adjusted for each amplifier.

The fourth stage is a single-ended amplifier with a 2N793 subminiature, high-gain silicon transistor. The multiplexer is direct-coupled to this stage.

Table 1 shows the specifications for a typical amplifier that was constructed.

One of the actual amplifiers used is shown in Fig. 3 before potting. The module shown also contains the multiplexing circuitry associated with this particular amplifier. To illustrate the reduction of size provided by the use of subminiature components and module construction, the original breadboard of the amplifier is shown in Fig. 4, without the multiplexing circuitry.

IV. THE MULTIPLEXER

In order to combine the outputs of the amplifiers into a single input to modulate the transmitter, a multiplexer is required. For this particular application, the multiplexer must be easily extendible, if desired, to more than four channels.

The two principal types of multiplexing are frequency multiplexing and time multiplexing. In the former, the modulated signals are combined into one wideband channel accommodating the entire range of frequencies involved. The carriers must be in the audio and superaudio ranges; rather bulky tuned circuits are required to generate these carriers. Also, a modulator is required for each channel, which adds to the complexity of the circuit. The system is easily extendible to more channels. However, the disadvantages of frequency multiplexing were considered to outweigh the advantages; thus it was decided to use time multiplexing.

Time multiplexing is the transmission of samples of information from several single channels through one communication system with different channel samples staggered in time.

Four types of time multiplexing systems were considered. These were

1) Pulse-code modulation (PCM)
2) Pulse-amplitude modulation (PAM)
3) Pulse-position modulation (PPM)
4) Pulse-duration modulation (PDM)

Of the four systems, PCM has the best signal-to-noise ratio.[16] This would be worth considering if it were not for the complexity of the circuitry needed for coding. Pulse-amplitude modulation suffers from the fact that information is contained in the amplitude of the pulses. Since most forms of noise produce amplitude changes in the signals, the signal-to-noise ratio suffers. These facts limit the choice to PPM and PCM.

In both PPM and PDM the information is contained in the position of the pulse edges, not in the amplitude of the pulse. This results in higher signal-to-noise ratios, at the expense of an increase in the bandwidth required, than can be obtained with amplitude-modulation systems. Pulse-duration modulation was chosen because of the simplicity of modulation and demodulation. This choice was made at the risk of cross talk, since the moment of switching on of a pulse is dependent on the slope of the trailing edge of the preceding pulse. This dependency becomes important after rf transmission through an FM system that slightly distorts the transmitted waveforms because of bandwidth limitations. This PDM system is easily extended to more channels. The four-channel system actually constructed was designed for a 2000-cps bandwidth for each channel. With the receiver employed, it was found empirically that this is the maximum single-channel bandwidth that is possible with four-channel PDM multiplexing. Thus if more channels are added, the bandwidths of all channels must be correspondingly reduced. This feature of the system is considered in more detail in Appendix A.

The output of the multiplexer is a staircase function whose steps vary in width, as illustrated in Fig. 5.

All edges of this waveform are used for the transmission of information. When

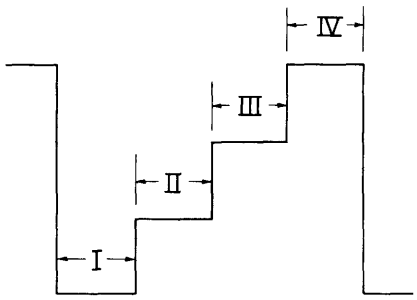

Fig. 5. Output waveform of multiplexer.

differentiated, all but one of the resulting pulses are of the same polarity, the odd pulse serving as a synchronization pulse, as well as an information pulse.

To generate this staircase function, one astable multivibrator and N-1 monostable multivibrators (N is number of channels; in this system, N=4) are connected in a ring, as illustrated in Fig. 6. All multivibrators produce pulses whose durations are proportional to the amplifier output signals associated with them.

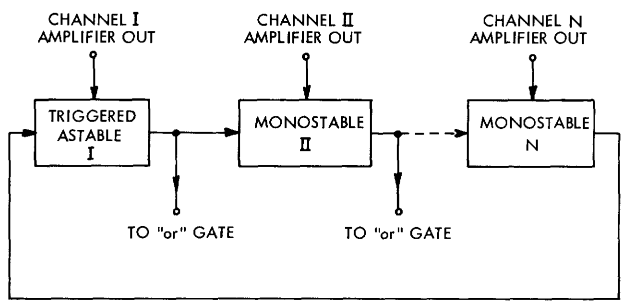

Fig. 6. Ring of multivibrators used in PDM multiplexer.

The purpose of the triggered astable multivibrator is to generate the initial starting pulse and to generate a restarting pulse if it should not receive a trigger pulse. In normal operation a trigger pulse arrives before it can generate its own pulse so that is behaves like a monostable multivibrator.

A major problem was to design a monostable multivibrator whose pulsewidth is linearly related to the amplitude of the input signal. A conventional pulse-duration modulated monostable multivibrator is shown in Fig. 7. V_{cc} is the supply voltage and v is the input modulating voltage.

The timing circuit for the monostable multivibrator is shown in Fig. 8.

Typical values that would be used in this system are v_o = 1.8 volts, V_{cc} = 2.5 volts, and v ranges from 1.3 to 2.3 volts.

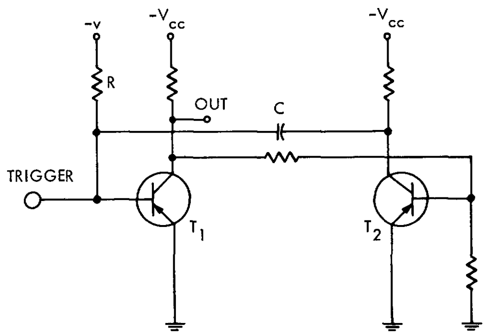

Fig. 7. A conventional pulse-duration modulated monostable multivibrator.

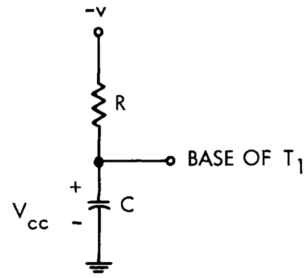

Fig. 8. Timing circuit for monostable multivibrator in Fig. 7.

A sample calculation shows that at the extremes, that is, with v = 1.3 volts or 2.3 volts, the nonlinearity is 19.4 per cent.[17]

To linearize the monostable multivibrator, a well-known method was employed. If the capacitor voltage at the start of the timing cycle is equal to the input signal v and a "constant-current source," I, is used to discharge the capacitor, the exponential nonlinearity in the example above is eliminated. The linearity, then, is restricted only by the "constant-current source," which, here, is $w = Cv/I$.

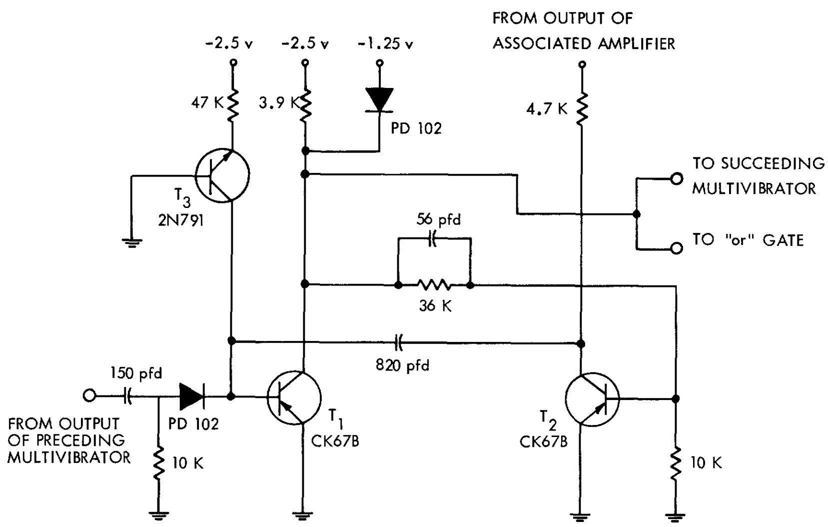

Fig. 9. Linear pulsewidth modulated monostable multivibrator.

A grounded-base transistor is a good approximation to a constant-current device and is shown in the actual pulsewidth modulated monostable multivibrator in Fig. 9.

A diode clamp to -1.25 volts on the collector of T_1 decreases the rise and fall time of the output pulse.

The triggered astable multivibrator is a slight modification of the monostable

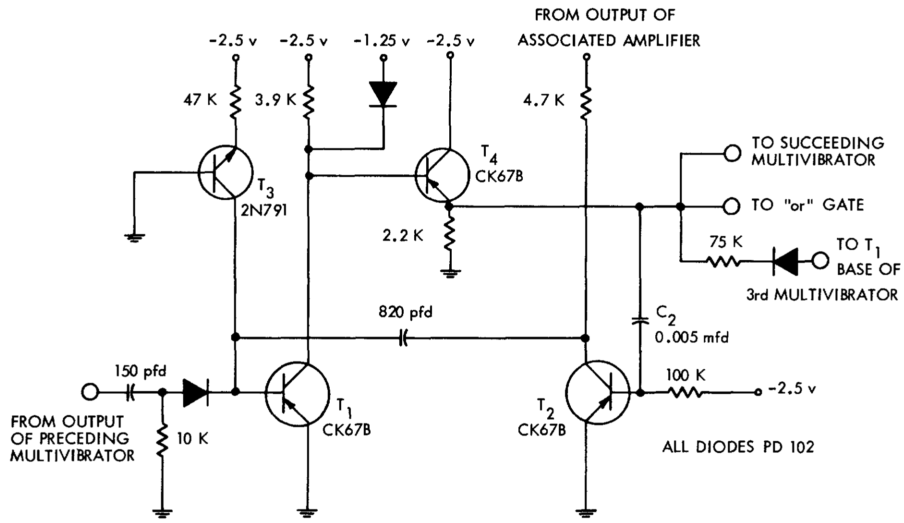

Fig. 10. Linear pulsewidth modulated triggered astable multivibrator.

multivibrator. The untriggered frequency of the astable multivibrator is 4000 cps, which is lower than the lowest possible frequency of the triggering pulses. The circuit for the astable multivibrator is shown in Fig. 10. The series resistor-diode combination on the output is connected to the T_1 base of the third monostable multivibrator. This connection guarantees that only one multivibrator is in its quasi-stable state at any given time.

The emitter follower on the collector of T_1 is used to prevent capacitor C_2 from increasing the fall time of the output pulse.

The outputs of the multivibrators go to an "or" gate to be combined into a single signal. Each is attenuated by a different amount to obtain the staircase function. The "or" gate and attenuators are shown in Fig. 11.

Note that only three of the four multivibrators are connected to the "or" gate. During the fourth pulse, no signal is applied and the staircase function assumes its least-negative value.

Each channel is sampled by its respective multivibrator at a mean frequency of 10,000 cps. The minimum and maximum possible sample frequencies that occur, if all channels have 250-μv peak-to-peak in-phase signals applied to the inputs of their respective amplifiers, are 7500 cps and 14,800 cps. From the Nyquist sampling theorem, this

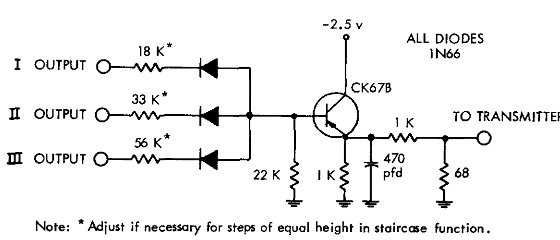

Fig. 11. Four-channel "or" gate and attenuators.

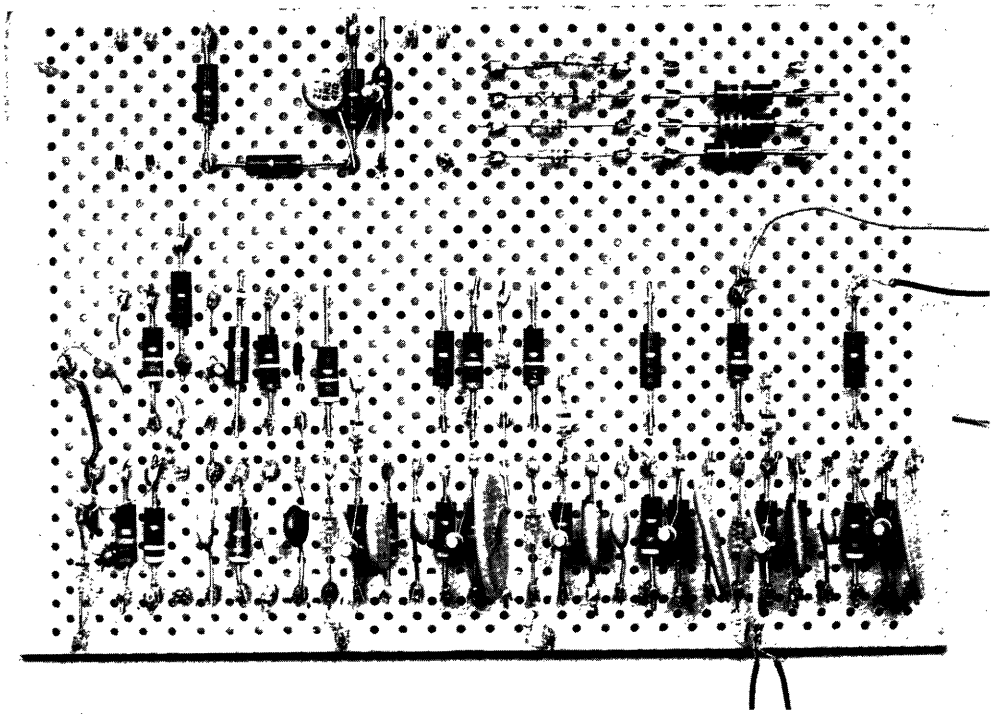

Fig. 12. Breadboard of multiplexer.

Table 2. Multiplexer characteristics.

Type	Pulse-duration modulation
Sampling frequency per channel	10,000 cps (mean) 14,800 cps (maximum) 7,500 cps (minimum)
Sensitivity	16-μsec pulsewidth change per volt
Linearity of pulsewidth change as a function of input-voltage change (dynamic input: 4 mv-1 volt peak-to-peak)	1 per cent (dc to 2000 cps)
Input impedance of typical input	4.7 ohms (minimum – immediately after sampling) 30 kohms (maximum – immediately before sampling)
Output impedance	68 ohms
D-C power requirements	-1.6 volts ± 20 per cent (from each amplifier) -2.5 volts at 3 ma

means that the maximum bandlimited signal that the minimum sampling rate will completely characterize is 3750 cps, a value greater than 2000 cps (the bandwidth of each channel), and thus it allows for practical filtering in the reconstruction of the waveform.

The multivibrators were constructed with their associated amplifiers in common modules. One of these amplifier-multivibrator modules is shown in Fig. 3. The breadboard of the entire multiplexer is shown in Fig. 12.

Table 2 gives the specifications for the multiplexer.

V. TRANSMITTER AND RECEIVER

Frequency modulation (FM) in the 88-108 Mc range is used for the rf information link. Operation in the VHF range was chosen because noise level is lower, circuit components are smaller, and relatively inexpensive receiving equipment is available. Frequency modulation is used because of its inherent advantage over amplitude modulation (AM) with respect to signal-to-noise ratio. The limiting system of the FM receiver also eliminates spurious AM signals that are due to changes in the distance between the transmitter and the receiving antennae which are caused by movements of the subject.

As Forsen has shown, a considerable amount of time can be devoted to the design of such a transmitter.[18] At this time there have been several adequate circuits developed[11,13,19,20]; thus it was decided to modify one of these for this application.

The transmitter that was modified is shown in Fig. 13 and was fully described by Thomas and Klein.[20] The 2N499 transistor performs the three functions of rf oscillation, frequency modulation, and amplification of the multiplexed input signal. Frequency modulation is accomplished by feedback-loop phase shift resulting from changes in the alpha cutoff frequency of T_1 under the control of the input signal.

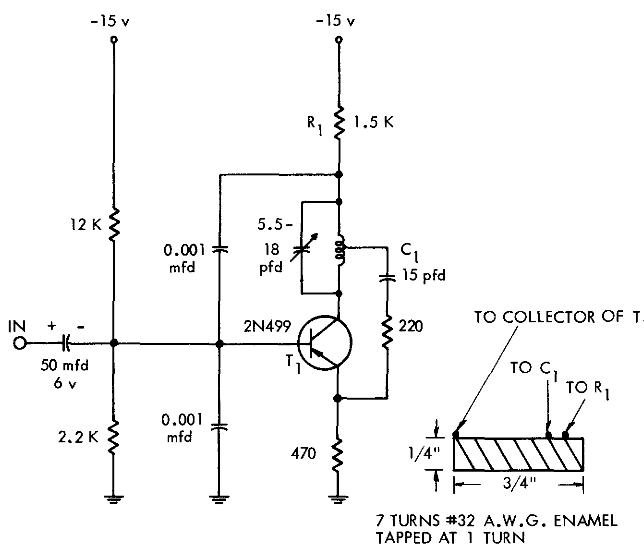

Fig. 13. Circuit diagram of FM transmitter and coil winding data.

Wideband frequency modulation provides a significant signal-to-noise improvement over narrow-band frequency modulation. The former requires that the modulation index β be greater than or equal to $\pi/2$.[16] For a sinusoidal modulating signal at frequency f_m

$$\beta = \frac{\Delta f}{f_m},$$

where Δf is the maximum frequency deviation away from the carrier frequency.

It was found empirically that the transmitter produces linear frequency modulation with 1000-cps sinusoidal input signals in the range 1-80 mv peak-to-peak at 68 ohms source impedance. A modified Harman Kardon Citation III receiver was used in this measurement (the modification is discussed with the receiver). At the maximum input level, that is, 80 mv, the frequency deviation Δf was 190 kc. Thus, for this measurement

$$\beta = 190.$$

Identical results were obtained by using a Boonton 202E FM signal generator instead of the transmitter. This result implies that the limiting factor in the FM link is the bandwidth of the receiver. For $\beta \gg \pi/2$, the bandwidth B of the FM signal[16] is approximately

$$B \doteq 2\Delta f = 380 \text{ kc}.$$

Since 380 kc is the largest bandwidth signal that can be passed through the receiver, the bandwidth of the receiver must also be approximately 380 kc.

The input signal to the transmitter is the step function described earlier and has frequency components up to 100 kc. A 100-kc lowpass filter was put at the output of the multiplexer.

An analysis[17] shows that the FM link is operated on the borderline between wideband and narrow-band frequency modulation. Since narrow-band frequency modulation has a poorer signal-to-noise ratio, it would be expected that noise in the FM link would be predominantly in the frequency band from 30 kc to 100 kc. This was found empirically to be true.

Table 3 shows the transmitter specifications.

Table 3. Transmitter specifications.

Frequency range	88-108 Mc (tunable)
Input for 100 per cent deviation	52 mv peak-to-peak[a]
Frequency response	±3 db, 10-100,000 cps
Power output	30 mw
Power requirements	-15 volts at 4 ma
Maximum range	200 feet

[a]Step function from multiplexer.

The receiver used is a Harman Kardon Citation III. This receiver has a rated sensitivity of 0.65 μv for 20 db of quieting and normally has a bandwidth of 200 kc at ±3 db. The modification mentioned previously was to increase the bandwidth to approximately

380 kc by retuning the first, second, and third intermediate frequency transformers, the limiter, and the discriminator transformers. This retuning was probably done at the expense of sensitivity; however, no attempt was made to measure this loss. The receiver also has excellent automatic frequency control (AFC) characteristics that enable it to track slow frequency changes in the transmitter which result from drift and capacity changes in the tank circuit as the subject approaches conducting objects.

VI. PULSE SHAPER AND DISTRIBUTOR

The staircase signal from the receiver usually contains noise and is somewhat distorted. The purpose of the pulse shaper and distributor is to remove as much of this noise as possible, to generate pulses that are identical in width to those generated at the outputs of the multivibrators in the multiplexer, and to distribute these pulses to their respective demodulators. A block diagram for this circuit is shown in Fig. 14.

The output of the highpass filter is positive and negative spikes for positive and negative level changes, respectively, of the input signal. This filter is followed by an amplifier and a noise limiter that has an adjustable threshold for passing the spikes. Since the noise output of the highpass filter is also in the form of spikes but generally of lower amplitude than the signal spikes, the threshold of the noise limiter can be set to pass the portion of the signal spikes with an amplitude greater than that of the noise spikes. The circuit of the filter, amplifier, and noise limiter is shown in Fig. 15.

The positive spikes from the noise limiter have smaller amplitude than the negative spikes. To correct this, they are amplified and inverted to negative spikes of approximately the same amplitude as the negative spikes from the noise limiter. This amplifier-inverter is shown in Fig. 16. These negative spikes trigger monostable B, and the negative spikes from the noise limiter trigger monostable A. Both monostable multivibrators are identical and produce pulses of 2-μsec duration at −6 volts for every input trigger. Emitter followers precede and follow each monostable multivibrator. The purpose of the emitter followers is to prevent loading of the input and output pulses of the multivibrators. One of the emitter follower-monostable multivibrator combinations is shown in Fig. 17.

The flip-flops are shown in Fig. 18. They have rise and fall times of less than 0.1 μsec with output levels of zero and −6 volts. A negative pulse of −6 volts with a duration of 0.5 μsec or longer is used as a trigger.

The logic of the distributor is very simple. A negative pulse from monostable A sets flip-flop I but applies a reset pulse to all other flip-flops. This pulse is always followed by a series of negative pulses (three for a four-channel system) from monostable B. These are applied to a reset input on each of the flip-flops (one flip-flop for each channel). The first negative pulse from monostable B then resets flip-flop I. However, the set output of flip-flop I connects to the set input of flip-flop II through highpass filter I (see Fig. 19) which has a time constant slightly greater than 2 μsec. Thus the reset pulse on flip-flop II from monostable B goes off, but the negative set pulse from highpass filter I is still present at the set input of flip-flop II. This causes flip-flop II to change to its set state. The second negative pulse from monostable B resets flip-flop II which, in turn, causes flip-flop III to change state. After the third negative pulse from monostable B has reset flip-flop III and set flip-flop IV, a negative pulse from monostable A, which sets flip-flop I and also resets flip-flop IV, starts the cycle over again.

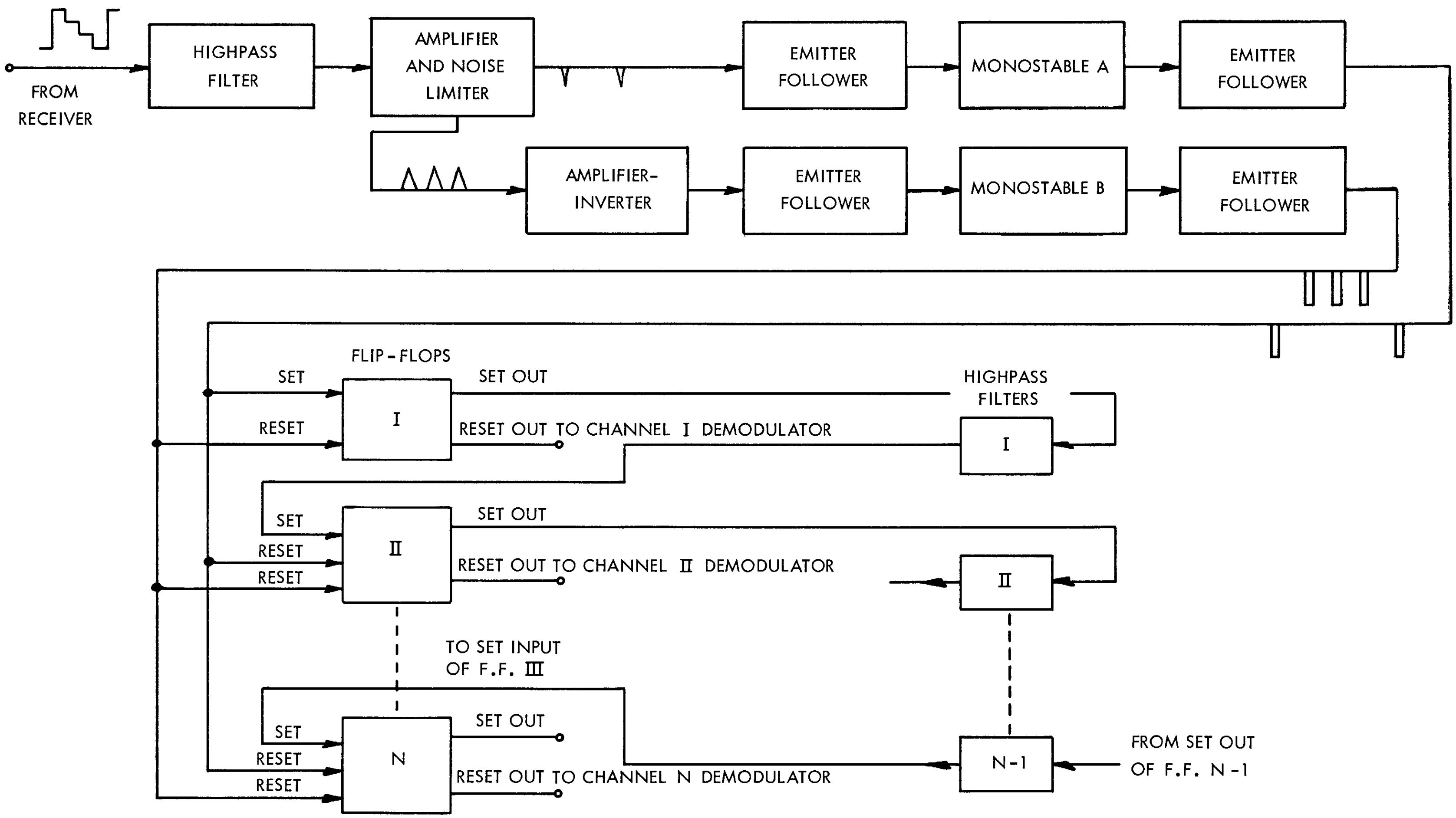

Fig. 14. Block diagram of pulse shaper and distributor.

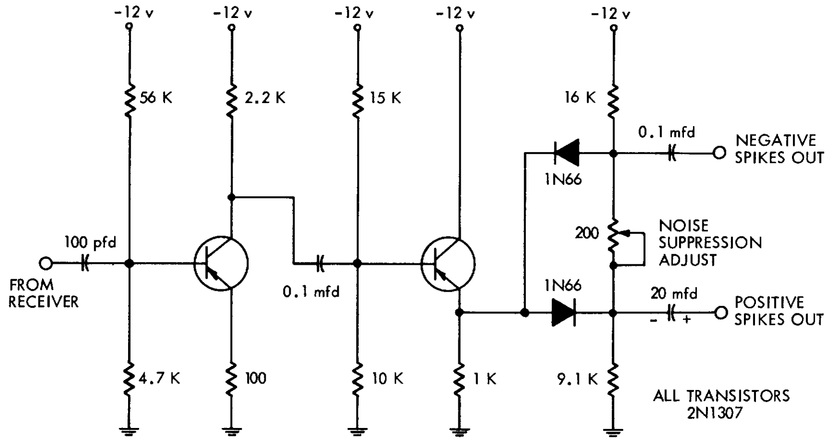

Fig. 15. Highpass filter and noise limiter.

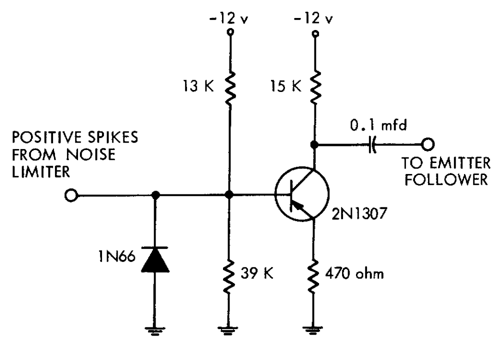

Fig. 16. Amplifier-inverter.

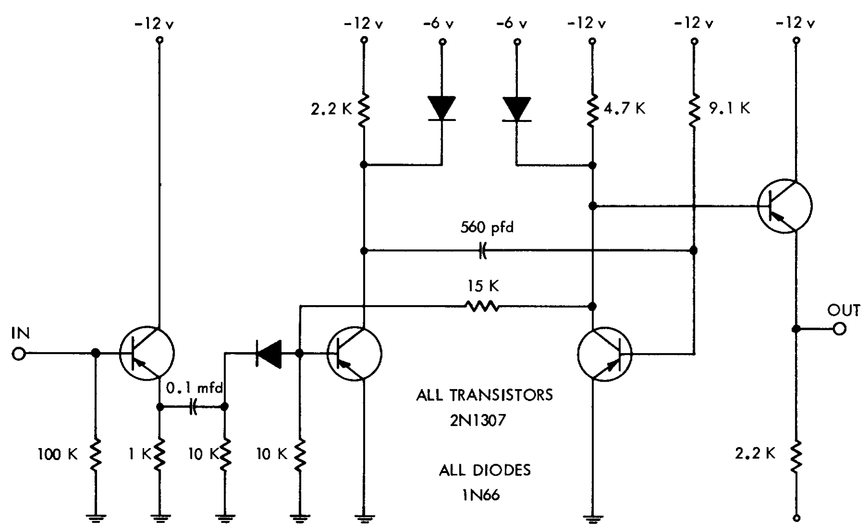

Fig. 17. Monostable multivibrator and associated emitter followers.

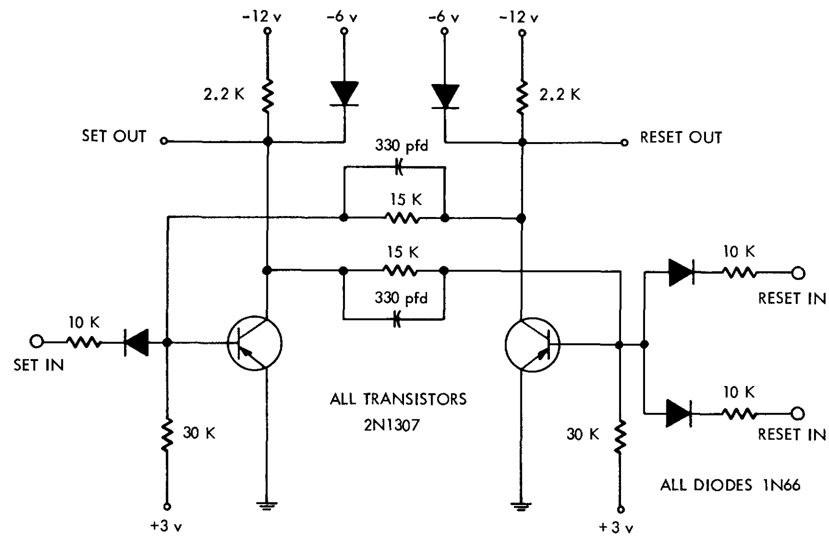

Fig. 18. Flip-flop.

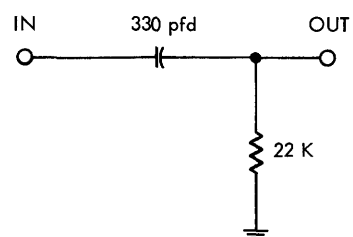

Fig. 19. Highpass filter.

The reset output of any particular flip-flop puts out a negative pulse of the same duration as its associated multivibrator in the multiplexer. These are then used to drive the demodulators.

As can be seen, to add more channels, one new flip-flop and one new highpass filter are all that are needed in the pulse shaper and distributor.

VII. THE DEMODULATOR

The outputs of the pulse shaper and distributor are negative pulses whose widths are modulated. Since the pulses have now been separated, the remaining problem is to convert the width-modulated pulses back to the original input signals. This is done by the demodulator, one of which is required for each channel. A block diagram of a demodulator is shown in Fig. 20, and a circuit diagram is shown in Fig. 21.

Transistor T_1 is connected in the grounded-base configuration and acts as a "constant-current source" for charging C_1. This charging occurs only during the presence of a negative pulse from the pulse distributor on the anode of D_1. The voltage across C_1 increases linearly to a voltage that is proportional to the width of the negative pulse, which is, in turn, proportional to the amplitude of the signal at the input of the channel. The capacitor C_1 is discharged through D_1 at the completion of the negative pulse. The charging voltage across C_1 is transferred through a Darlington circuit consisting of T_2 and T_3 to capacitor C_2. The Darlington circuit has a high input impedance so as not to affect the linearity of the voltage build-up on C_1. C_1 discharges rapidly, but the voltage across C_2 discharges more slowly through R_1 with a time constant of 15 μsec.

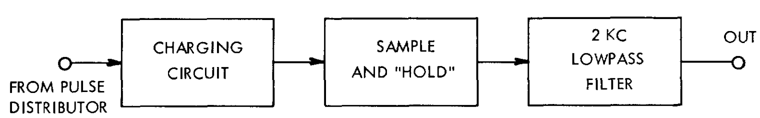

Fig. 20. Block diagram of a demodulator.

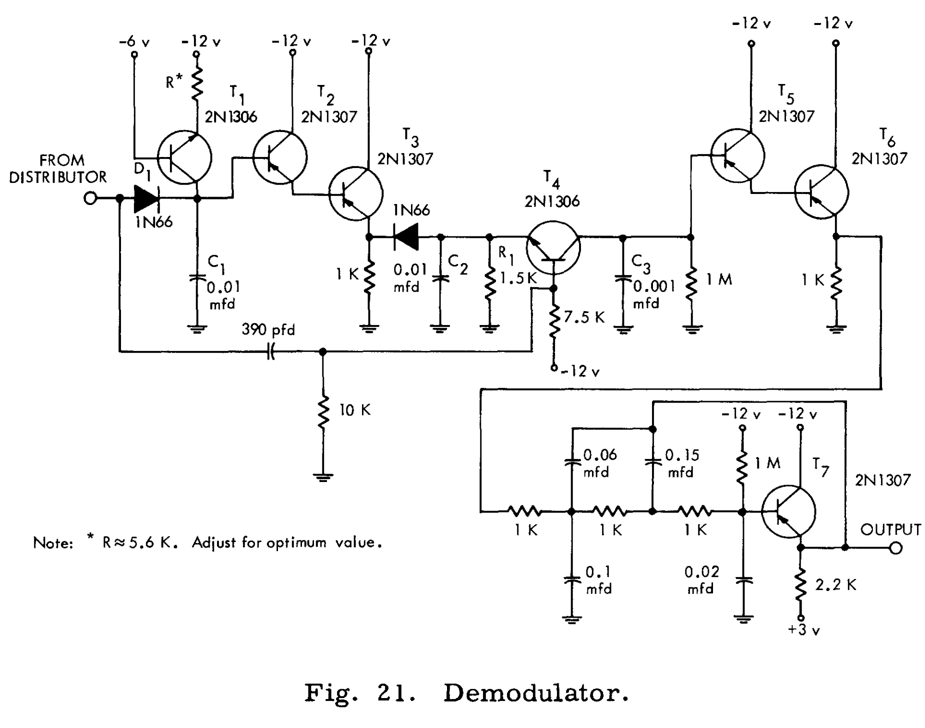

Fig. 21. Demodulator.

Immediately after C_1 and C_2 reach their peak value, a 1-μsec positive pulse that is derived from the positive-going pulse on the input to the demodulator is applied to the base of T_4, which is normally cut off. This causes T_4 to conduct for 1 μsec with the result that the peak voltage on C_2 is applied to C_3, which holds this value until the next positive pulse is applied to T_4. The 15-μsec time constant guarantees that the voltage change across C_2 is negligible during the 1-μsec sampling period. The second Darlington circuit, consisting of T_5 and T_6, is used as a high-impedance input, low-impedance output device so that the voltage across C_3 does not change appreciably between sampling periods.

A third-order Butterworth lowpass filter connected to the output of the Darlington circuit has a 3-db fall-off point and is particularly effective in removing harmonic components about the sampling frequency. (A more complete analysis of the demodulator has been given.[17])

VIII. THE COMPLETE SYSTEM

The telemetry package carried by the subject is shown in its actual size with one side removed in Fig. 22. The case was machined from 0.0938-in. nylon sheet with 0.0625-in. partitions. The weight of the complete package, including batteries, is 212 grams.

The decoding electronics that are not carried by the subject have been built on plug-in cards, as shown in Fig. 23. The top photograph shows the flip-flops and their associated highpass filters; the middle, two of the demodulators; and the bottom, the noise limiter, the monostables, and their associated amplifiers and emitter followers. Not shown is one demodulator card that is identical to the one shown in the middle photograph of Fig. 23. The receiver is a Harman Kardon Citation III

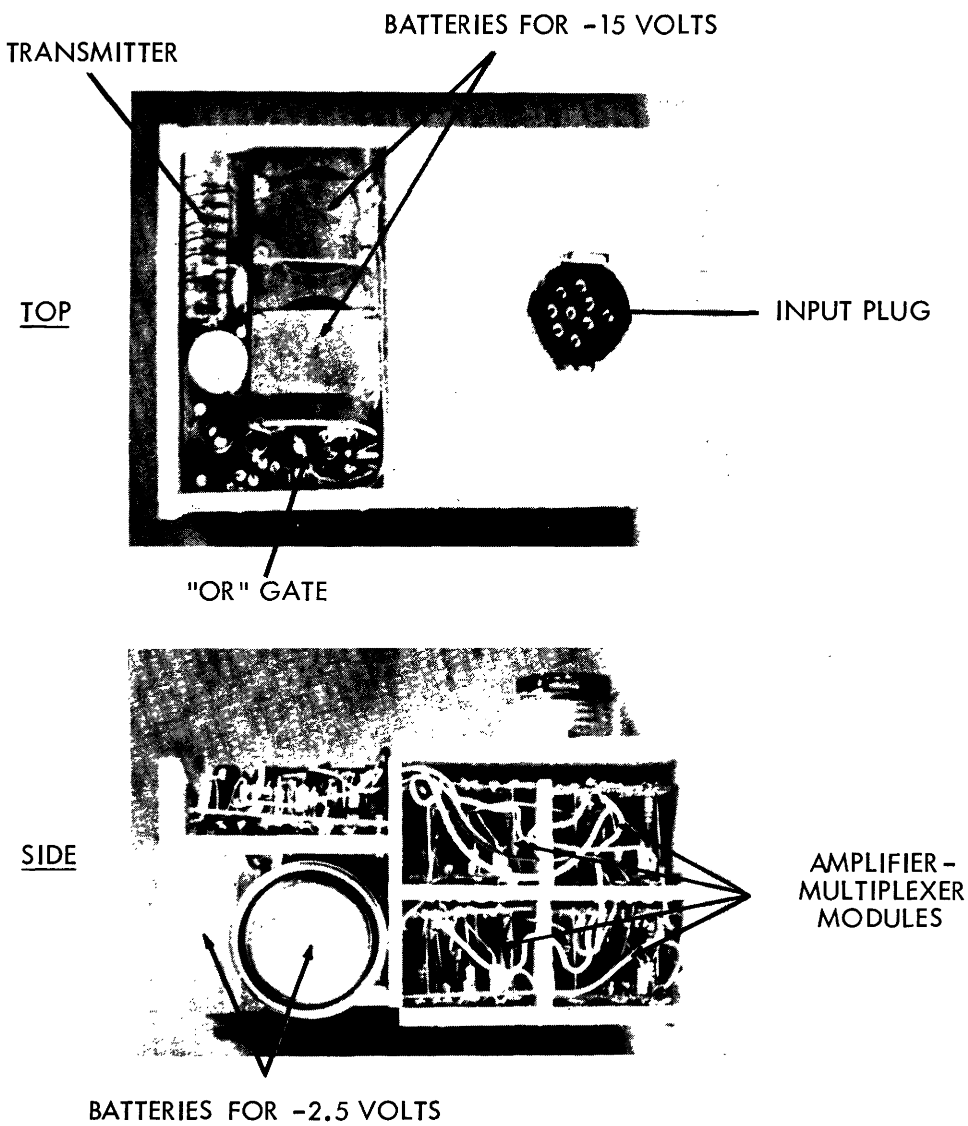

Fig. 22. Telemetry package with one side removed (actual size).

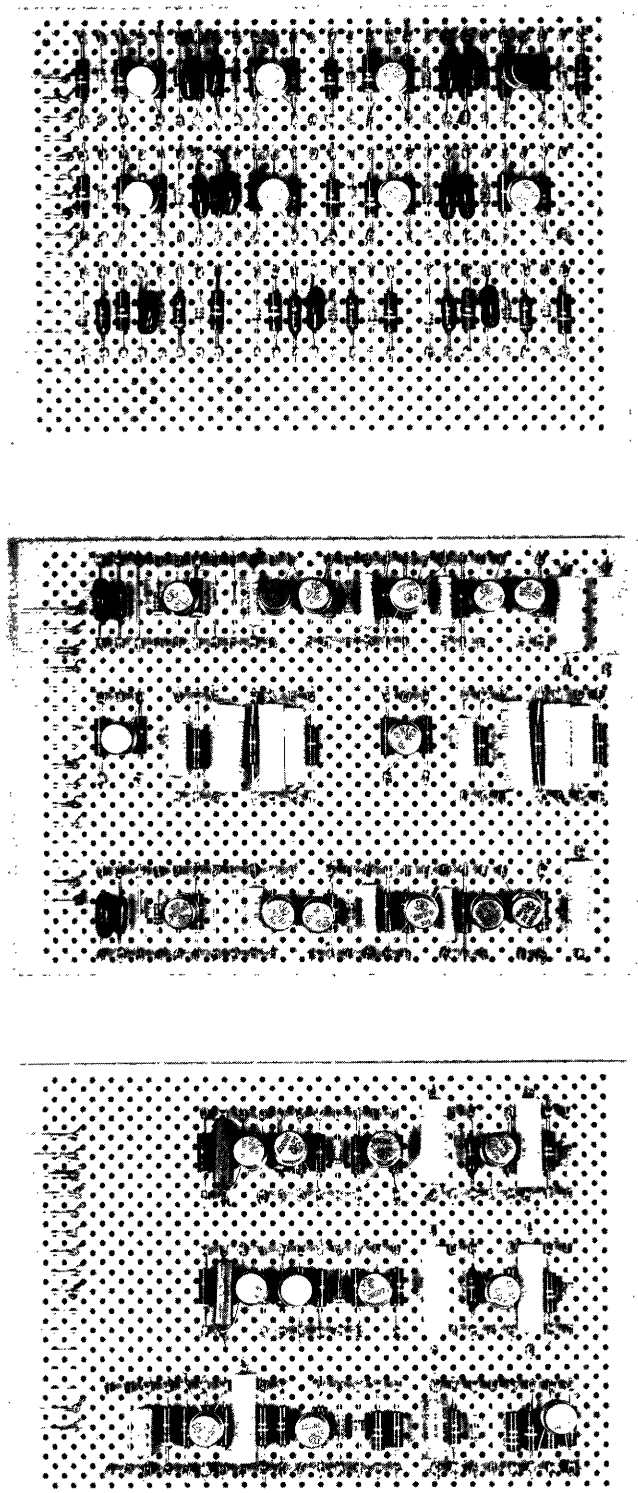

Fig. 23. Decoder plug-in cards.

Table 4. Specifications of complete telemetry system.

Number of channels	4
Type of multiplexing	Pulse duration
Frequency response	±3 db, 0.7-2000 cps (each channel)
Dynamic range of inputs	5-250 μv (each channel)
Noise (referred to input)	5 μv peak-to-peak (0.7-2000 cps)
Cross talk between channels	Less than 3 per cent
Transmitter radiofrequency	88-108 Mc (tunable)
Operating range	200 feet
Battery life	12 hours between recharging
Power requirements	−2.5 volts at 6 ma, −15 volts at 4 ma } Telemetry package −12 volts at 200 ma, −6 volts at 15 ma, +3 volts at 13 ma } Decoder

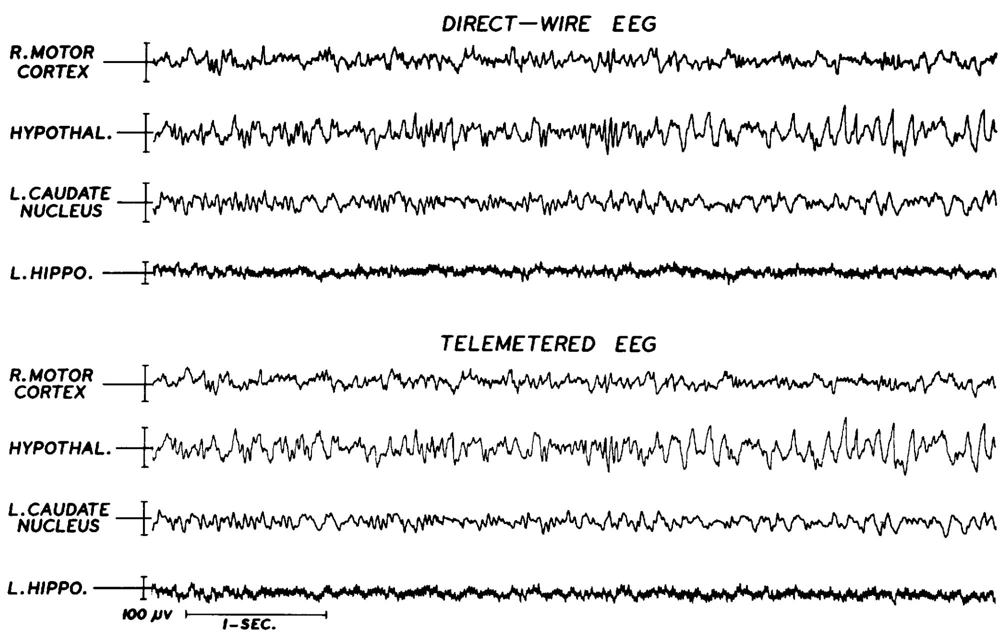

Fig. 24. Direct-wire and telemetered EEG from a dog.

modified as described in Section V.

The complete system has been tested in breadboard form and has the specifications given in Table 4.

Direct-wire versus telemetered EEG signals are shown in Fig. 24. These records were obtained simultaneously from an alert, freely moving dog. The dog has a defective electrode in the left hippocampus, as illustrated in Fig. 24 by the 60-cps ripple in both the direct and telemetered records from this region. An Offner Type R dynograph was used for the direct recording and the telemetry write-out.

APPENDIX A

EXTENSION OF THE SYSTEM TO MORE THAN FOUR CHANNELS

Of primary concern in extending the system to more channels is the fact that the sampling theorem must be satisfied. This imposes the requirement that if a signal has an energy spectrum with its highest frequency components at some frequency f_o, then it must be sampled at a frequency $f_s \geq 2f_o$ in order to recover the signal.

As an example, the changes that are necessary to extend the described four-channel system to eight channels will be discussed. Obviously, four more differential amplifiers must be added. These can be identical to the four that are already present. Since each amplifier draws 0.6 ma, the total increase in current drain will be 2.4 ma. Each new channel requires a pulse-duration modulated monostable multivibrator that is identical to those in the present system. These are simply inserted in the multivibrator ring as shown in Fig. 6. A current drain of 0.7 ma is added for each new multivibrator.

Since the total period between sampling of each channel is now doubled, the sampling frequency has been halved. Because of the practical aspects of filtering and the fact that the mean sample frequency of each channel has now been reduced from 10,000 cps to 5000 cps, the bandwidth of each channel must be reduced from 2000 cps to 1000 cps, that is, the EEG signals of interest must have only negligible frequency components higher than 1000 cps. This restriction, expressed for N channels, where B is the bandwidth of each channel, is

$$B = \frac{8000}{N}.$$

As can be seen, an eighty-channel system theoretically could be built in which each channel has a 100-cps bandwidth; however, it will be shown that noise may limit the number of possible channels. For each channel added a new input must be added to the "or" gate (Fig. 11), and all of the attenuators must be adjusted so that equal level changes occur in the staircase function. The transmitter and receiver remain unchanged.

Since each level of the step function will be reduced correspondingly, the positive pulses obtained at the output of the noise limiter (Fig. 15) in the pulse shaper will also be reduced. This reduction requires that the gain of the amplifier-inverter (Fig. 16) be increased so that its output pulses are approximately equal in amplitude to the amplitude of the negative pulses out of the noise limiter.

The pulse distributor requires only that a new flip-flop and a new highpass filter be added for each channel added. These should be added in the manner illustrated in Fig. 14.

The final change is to add a new demodulator for each channel added. Since the last stage of the demodulator is a lowpass Butterworth filter with a 3-db point at the bandpass of the channel, all Butterworth filters in all of the demodulators should be frequency-scaled to the new bandpass. In the eight-channel example given, this would involve doubling the capacity of all capacitors in the Butterworth filters.

Problems with noise may limit the number of channels that can be added. In the pulse shaper, the output of the highpass filter contains both signal pulses and noise pulses. As more channels are added, the signal pulses decrease in amplitude, but the noise pulses are unaffected. Thus a point may be reached at which the noise limiter cannot discriminate between noise and signal pulses. This must be determined empirically, since the amount of noise is dependent on many variables, for example, radiofrequency used and atmospheric conditions.

APPENDIX B

CONSTRUCTION CONSIDERATIONS

As mentioned in Section III, the transistors for the differential amplifiers must be matched in pairs. The first stages use an FSP-2, which is a dual differential transistor already matched and packaged in a single TO-5 can.

The data for matching the other two pairs of transistors in the amplifier are given in Table 5.

Table 5. Data for matching transistor pairs for the differential amplifier.

Stage	Transistor	Bias at 24°C		β		I_{ceo} (amps) at given V_{CE}	I_{ceo}
		V_{CE} (volts)	I_C (amps)	(greater than)	(matched within)	(less than)	(matched within)
II	CK22B	1.00	100	30	5%	30	5%
III	CK67B	0.76	92	46	5%	50	5%

The subminiature components used are

- Resistors: Ohmite, 0.1 watt, 10 per cent composition carbon
- Electrolytic capacitors: Mallory TNT series
- Ceramic capacitors: Glenite 601 series
- Microdiodes: Pacific PD 102
- Transistors (except FSP-2 and 2N499): Raytheon CK22B, CK67B, 2N791, 2N793
- Differential transistor: Fairchild FSP-2
- Transmitter transistor: Philco 2N499
- Rechargeable batteries: Burgess CD-2 and CD-3

The case is machined from 3/32-in. nylon sheet. The printed circuit boards for the amplifier-multiplexer modules are compatible with both the triggered astable circuit and the monostable circuits discussed in Section IV, and are 1/32-in. epoxy G-10.

Acknowledgment

Thanks are due to A. Kamp for details of his systems and suggestions for the solution of specific problems. The amplifier circuit and parts of the demodulator circuit are adaptations of his original designs.

I would like to thank Professor Ronald Melzack for providing the initial stimulus for the system and his continued interest in the use of telemetry.

I would also like to thank Paul Donahue for his ideas concerning packaging and miniature components, and Stephen K. Burns for his assistance in obtaining components and his interest.

Special thanks are extended to Professor Moïse H. Goldstein, Jr. for technical criticism and suggestions and assistance with preparation of the manuscript.

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