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Solid State Electronics Laboratory

Engineering Design Center

Case Western Reserve University

by

W. H. Ko

E. Yon

D. Ramseth

W. Thompson

R. Antinone

TABLE OF CONTENTS

LIST OF FIGURES

LIST OF TABLES

SECTION I

SUMMARY

The purpose of this project is to study and apply advanced electronic technology to the design and development of an integrated circuit, multi-channel telemetry system for bio-medical applications. This system should be implantable and be capable of telemetering a wide range of physiological signals. The system should be flexible, allowing specific telemetering requirements to be met within the system design framework. The circuitry should be modularized to form building blocks from which a variety of systems can readily be formed.

A time division multiplex system was chosen for development because the transmitter circuitry required for this type of system is compatible with both hybrid and monolithic integrated circuit technology. Also, this type of system allows the possibility of reducing power drain by gating the power to some circuitry so that it is only on when it is required for the operation of the transmitter. A one hundred percent (100%) duty cycle pulse amplitude modulation multiplex system was chosen because of circuit simplicity along with adequate signal recovery accuracy.

A PAM-FM (pulse amplitude modulation-frequency modulation) scheme of modulation was chosen for initial development of the system since FM receiving equipment with sufficient RF bandwidth

for this type of modulation is readily available, and therefore the design of a receiving system would not be necessary at this time.

A discrete component, single channel FM-FM transmitter was constructed and implanted in a dog to test the RF section of the transmitter for the proposed system. Also, evaluation of water proofing and packaging was made to determine the extent of the interaction between the dog and the electronics of the transmitter.

A transmitter with four strain gauge channels was constructed using discrete microminiature components. This transmitter was implanted in a dog to evaluate the reliability and accuracy of the developed system.

In parallel with the effort on the transmitter was the development of a demodulator for the system. The demodulator, as developed, is compatible with a system of up to eight (8) information channels. Expansion of the demodulator to accommodate more channels can be readily accomplished. Some complexity was allowed in the demodulator circuitry to allow the use of simpler circuitry in the transmitter. However, all of the circuitry was kept as simple and as automatic as possible to eliminate the need for a trained electronics technician to operate the system.

The transmitter circuitry was reduced to modules, each of which it was possible to build in a multilead flat pack using hybrid integrated circuit technology. A four channel transmitter using flat pack circuitry which was essentially the same as that used in the discrete component transmitter was built and tested.

Modification of some of the transmitter circuitry allowed more of the previously defined modules to be placed in each flat pack resulting in an overall reduction in volume. Also, a reduction in the power drain of the transmitter resulted from these modifications. A four channel transmitter using the new circuitry was built and tested.

Circuitry was breadboarded to allow the telemetering of EEG and EKG signal information. An EKG signal obtained from surface electrodes on a human, and an EEG signal from electrodes implanted in the brain of a cat were successfully telemetered using the breadboarded transmitter. The circuitry for the EKG and EEG signals has not as yet been put in flat pack form.

A portion of the demodulator was redesigned using commercially available MOS integrated circuitry. The new circuitry allows for shorter sampling periods than were previously possible, and will therefore allow other requirements on the transmitter and receiver portion of the system to be relaxed.

SECTION II

DESIGN CONSIDERATIONS

Physiological studies often require simultaneous monitoring of several variables. A wireless measurement is imperative in many instances to eliminate the psychological artifact present in the behavior of a subject hard-wired to monitoring and recording equipment. In chronic studies an implanted transmitter is especially desirable. This offers freedom from infection from lesions in the skin for sensor leads as well as elimination of psychological artifact.

Appendix I gives a survey of some available multi-channel physiological telemetry systems. These designs are not applicable to a general-purpose, wide-bandwidth, implantable physiological telemetry system. They are limited by considerations of physical size, power supply lifetimes, and bandwidth.

The signal handling capabilities must be versatile, allowing for different numbers of channels and different bandwidth characteristics, as required by a particular system application. Optimization for a given set of signals must be possible.

The transmitter unit is to be miniaturized and physiologically inert so that it can be implanted in a small animal, such as a dog or possibly a monkey. The construction of the transmitter shall allow for the fabrication of large quantities of units. This implies the fabrication should not be critical or tedious so that performance will not vary from one transmitter package to another.

The circuitry of the transmitter must be simple and non-critical because of the reliability required by the inaccessible location of this unit in operation.

The completed system must be clinically useful. The system must be easily operated and yield dependable results. The operation of the system should require nothing more than the turning on and off of the transmitter and receiver units and recording the output of the system with the proper equipment.

Since batteries will be used for power, the circuitry should require a minimum of power from its power source and still maintain reliable operation. Even if rechargeable batteries are used, it is desirable to recharge as infrequently as possible.

The RF link for the system must be capable of transmitting the signal from the implant to the external monitoring system. The noise level of the entire link including transmitter, receiver and demodulator should be comparable with that of a hard-wire system.

An RF transmission link range of 10 to 20 feet should be adequate for the system. Most monitoring applications will permit close proximity of the subject and a receiving antenna, usually in an open area or open room.

Signal processing and telemetering is the primary function of the system. The design must have the capacity to telemeter the desired physiological signals. Appendix II gives a discussion and tabulation of physiological signals which have been telemetered in various studies. This study shows that signals from essentially DC up to 500 Hz are clinically monitored. EMG signals have frequency components up to 10^5 Hz, but clinical monitoring only makes use of those up to 500 Hz. Bandwidths to about 3 kHz are sometimes desirable for EMG studies. An accuracy of one percent (1%) of full scale is adequate for most studies.

Eight-channel recorders are standard for hard-wire monitoring systems. A total information bandwidth of 20 kHz would allow 2.5 kHz response for each of eight channels. Fewer channels could have a wider bandwidth for each, with a total adding up to 20 kHz. This total bandwidth possibility is the goal for the system design framework.

Some distinct advantages are offered, in the circuit design, by the implant environment. The transmitter is completely enclosed in a conductive medium. This provides very good shielding properties. Interference from external sources is isolated from the circuitry within the implanted unit.

The implant environment is also very closely temperature regulated. The transmitter will see less than $5^{\circ}C$ temperature change in operation and less than $20^{\circ}C$ from room temperature tests to implanted operation. The transmitter unit must maintain operating characteristics constant for a maximum temperature change of $20^{\circ}C$. This allows much freedom in the circuit design.

Integrated-circuit construction techniques are required by considerations of size, ease of fabrication, and repeatable performance from unit to unit. In this technique, functional units of circuitry are fabricated in individual sealed packages. These units are tested functionally and accepted or rejected as a whole unit. These units are then building blocks used in constructing the larger system. Constructing a complete transmitter then entails selecting and interconnecting the proper complement of pre-tested units. This procedure yields completed units with similar characteristics.

Initial evaluation of the circuitry should be done using hybrid integrated circuit construction. However, the circuitry shall be such as to allow later fabrication in monolithic integrated circuit form. Hybrid integrated circuit construction uses chips (also called dice or pellets) containing individual components. The chips are bonded to patterns on ceramic substrates. Interconnecting wiring is done with .001 inch diameter aluminum or gold wire. The completed circuit is sealed in a multi-lead package. This procedure is discussed and examples shown in Appendix III.

Available circuit components are limited by this construction technique. Inductances are not available. Capacitors, up to about 500 pf, are available in chip form. Resistances are available up to 50 kΩ per chip. A wide selection of transistors is available. Active devices are cheaper and more plentiful than passive devices. As a consequence of these limitations, circuit designs should incorporate large numbers of transistors, some low-value resistors, very few capacitors, and no inductors.

The tolerances of these chip components are quite wide. Specific values can be selected prior to fabrication, but their characteristics are affected irreversibly by the temperatures involved in bonding. These changes are not predictable or repeatable quantitatively. The only parameter which remains quite stable is resistance ratios in a single chip. These are determined by geometry and not affected by thermally-induced changes in diffusions, etc. Thus, circuit performances should be independent of components' parameters other than resistance ratios.

The receiving system must be completely automatic in operation in order to be clinically useful. Standard circuitry may be used. Complexity is allowed in the receiving system. Sophistication is to be used here to add reliability to the overall telemetry system.

SECTION III

TRANSMITTER DESIGN PARAMETERS

The transmitter design involves the selection of methods of multiplexing, of RF transmissions, and of supplying power to the circuitry of the implanted unit. These designs must be mutually compatible and meet the requirements set forth in Section II.

3.1 Multiplexing format

The multiplexing format chosen must accommodate the information bandwidth required and be instrumented in a size compatible with being implanted. It must be flexible, allowing freedom in numbers of channels, and capable of being optimized for a particular application. The circuitry necessary to instrument the format must be simple, use a minimum of components, and draw a minimum of power. The signal-handling capabilities, from Section II, are:

a) handle up to eight information channels

b) handle a total information bandwidth of 20 kHz

c) provide accuracies of one percent (1%) of full scale

Multiplexing is accomplished by frequency-division or by time-division. Frequency-division multiplexing uses subcarrier oscillators of frequencies up to about 200 kHz. The modulation methods for the subcarriers are:

a) AM - amplitude modulation

b) FM - frequency modulation

Time division multiplexing involves sampling the information channels and routing the sample pulses in sequence through the common communications link. The pulses are modulated by the information in the following forms:

a) PAM - pulse-amplitude modulation

b) PDM - pulse-duration modulation

c) PPM - pulse-position modulation

d) PCM - pulse-code modulation

A multiplexed signal from one of these methods is used to modulate the amplitude, phase, or frequency of an RF carrier. This gives, for example, an FM/FM telemetry system with frequency-modulated carrier and subcarriers.

3.1.1 Component and fabrication limitations

Frequency division multiplexed systems employ a low-frequency oscillator in each channel. These oscillators require the use of bulky components-capacitors and sometimes inductors. These oscillators require power continuously. In time division multiplexed systems power is mandatory only for the circuitry pertaining to the channel being sampled at any given instant or time period. Circuitry which is not processing information at a given instant may be turned off, if designed for this mode of operation.

The subcarrier oscillators in a frequency multiplexed system must have close tolerances on their center frequencies, if they are not to interfere with each other. This requires careful adjustment of each oscillator. Also, each channel takes a different frequency oscillator. In time-division multiplexed systems similar circuits are used from channel to channel.

These considerations recommend time-division over frequency-division in a miniaturized transmitter. Three basic categories of time-division-multiplexed systems are used. All sample signals obtaining pulses whose amplitudes are related to the analog input. In PAM the samples are transmitted, without conversion, over the RF link. Higher accuracies are obtained by converting pulse amplitudes into times as pulse durations (PDM), pulse positions (PPM), or pulse rates (PRM). Highest accuracies are obtained by converting amplitudes to digitally coded forms (PCM).

PAM is the simplest, requiring the least circuitry. The other methods require the same circuitry as PAM plus additional circuitry for conversion of pulse amplitudes to other codes. PCM requires very elaborate and power-consuming circuitry. Conversion circuitry for PDM is not as elaborate as for PCM, but is still too extensive and power consuming for an implantable transmitter design. Therefore, PAM appears the best modulation for the time-division multiplexed system judged by power requirements, component requirements, and ease of fabrication. It must, however, have the bandwidth and accuracy capabilities required for the system.

3.1.2 Bandwidth and Accuracies

A theoretical study has been reported which evaluates PAM/FM, PCM/FM, PDM/FM, and FM/FM for systems with varying degrees of accuracy[1]. For accuracies of two percent (2%) of full scale in the signal channels, PAM/FM is the best. Accuracies of 0.5 percent and better require PCM. PAM/FM, with 100 percent duty cycle pulses, requires the least video bandwidth and transmitted RF power. Details of this study are discussed in Section IV.

Minimum bandwidth for a given task is important in a wide bandwidth system. Larger RF bandwidths require more transmitted power and special receiving equipment.

Several modulation methods, for systems telemetering the same information, have been compared on the basis of RF bandwidth and signal-to-noise at the receiver input[2]. Table 3.1 gives the results quoted.

3.1.3 Multiplexing format selection

PAM was chosen, on the basis of the results of the two preceding sections, for the multiplexing method to be used in the transmitter. All the considerations investigated indicate PAM best.

It requires the least components, no large components, the least RF bandwidth and transmitted power, and offers the greatest ease of fabrication.

3.2 The RF transmission link

The choice of PAM places some requirements on the RF link. One of the most important is linearity. Analog information is present in each section of the telemetry channel and nonlinearities in any section will add into the overall transfer function for the channel. Since one-percent accuracy is required, the RF link must possess linearity better than one percent.

Gain stability and DC-level stability must be possessed by the complete communications link. Gain stability is necessary to maintain calibration accuracies. DC-level stability is necessary to maintain frequency response to DC. These requirements could be placed on the RF link but would be difficult to instrument. They can be met by including information in the composite PAM waveform for the demultiplexing unit to interpret and utilize to reinsert DC levels and standardize gain.

The RF link must meet the following requirements, based on the preceding discussion and on the discussions in Section II.

a) linearity better than one percent,

b) ease of fabrication in miniature form,

c) sufficient power output for transmission over a distance of 20 feet to give a 40 db signal-to-noise ratio at the output of the receiver.

d) accommodate video bandwidth sufficient to telemeter a PAM waveform with a 20 kHz total information bandwidth for the signal channels.

There are three modulation methods for the RF link which offer excellent linearity, use simple circuitry, and can be fabricated in a form compatible with an implanted unit. These are frequency-modulation (FM), double-sideband amplitude modulation (DSB-AM) with a suppressed carrier, and ordinary amplitude modulation (AM). DSB-AM requires two RF coils while FM and AM require only one. Otherwise, the components and volumes required are similar.

A theoretical study[3] has been made of modulation methods applicable to a wide bandwidth physiological telemetry system. This study reports that DSB AM with a suppressed carrier, as obtained from a balanced modulator, is the optimum method. It offers:

a) Better noise and bandwidth possibilities than FM

b) Lower power requirements

c) Can be crystal-controlled

FM obtains noise improvement over AM by using an RF bandwidth much wider than the information bandwidth. As information bandwidth increases and RF bandwidth remains constant, this noise advantage becomes lost. This occurs as frequency responses of the communications link, called video bandwidths, of 200 kHz or higher are necessary. In addition, DSB AM detectors can cancel any noise at the output appearing in only one sideband.

DSB AM, with suppressed carrier, transmits power in the information-bearing sidebands only. This requires only one third of the power of conventional AM. Crystal-control of the carrier frequency is also possible with DSB AM. This is desirable in a system used by non-engineering personnel. Suitable crystals are currently being packaged small enough for an implanted transmitter. The crystal package would occupy a volume equal to the remainder of the RF stage.

The circuitry for DSB AM, however, is very critical, since it requires more than one RF coil, and, in order to obtain the power advantage over ordinary AM, it requires balancing in the RF circuitry. The critical nature of this circuitry and the difficulties involved in fabrication to obtain proper operation led to the choice of FM as the type of RF link.

Some work was done with ordinary AM, but problems were encountered here with crystal activity. At the low power levels which were desired, operation and adjustment were very difficult without the use of multiple coil RF configurations.

FM circuitry is relatively simple and proven to be reliable, so this was the natural choice for the development of the system.

TABLE 3.1

RF BANDWIDTH AND SIGNAL-TO-NOISE RATIO REQUIREMENTS FOR DIFFERENT MODULATION TECHNIQUES

Modulation Method	RF Bandwidth, kHz	SNR Required, dB
FM/FM	300	9
FM/PM	300	9
PAM/FM/FM	300	9
PDM/FM/FM	300	9
PDM/FM	50	9
PCM/FM	12.5-1500	13
FM/AM	up to 150	14
PAM/FM	30	9

SECTION IV

OVERALL SYSTEM DESIGN PARAMETERS

Several parameters must be determined in finalizing a PAM-FM system for a particular application. These are chosen to satisfy the requirements of bandwidths and accuracies of the several information channels. These parameters are:

a) Sampling ratio

b) Duty cycle of sample pulses

c) Channel, sync, and AGC pulse levels

d) Video bandwidth

e) RF carrier bandwidth

f) Peak RF carrier deviation and modulation index

g) Filters and bandwidth limits

A system design has been reported[4] for a PAM-FM telemetry system with a 2% accuracy. The design is outlined in Table 4.1. Theoretical results were compared with those obtained from an experimental system. This design provides guidelines for selection of parameters in the system design framework being developed by this project.

4.1. Sampling Ratio

Sampling ratio is the relationship between the sampling frequency of a channel and the maximum frequency-component telemetered by the channel. The "sampling theorem", in its simplest form, states that the sampling frequency must be at least twice the highest frequency in the waveform being sampled for the waveform to be reconstructed in its original form[5].

Frequency components present, higher than one-half the sampling frequency, are folded about the sampling frequency and its harmonics, appearing below one-half the sampling frequency as errors. Noise in the communications link between sampling and desampling gates also appears in the channel output as errors. This error generation is known as aliasing.

Sampling errors, including aliasing, are a function of filtering before sampling, sampling ratio, and filtering after desampling (interpolation). Errors under one percent require a minimum sampling ratio of four with a filter before sampling of at least 36 dB per octave rolloff. Interpolation filtering of 36dB per octave, with a bandwidth of 1.5 times the presampling filter, is required[6].

4.2 Duty Cycle of Sample Pulses

The duty cycle of a sample pulse is the ratio of the length of the pulse to the length of the time period for that channel. A 50 percent duty cycle is generally used. This gives sample and zero level pulses of equal length for each channel.

A duty cycle of 100 percent requires a smaller communications link or video bandwidth than any other duty cycle. A 50 percent duty cycle requires twice the bandwidth of that of 100 percent. Other duty-cycles require still greater bandwidths.

There are synchronization problems with a format using a duty cycle of 100 percent. Synchronization on each channel-change time is possible by differentiating and full-wave rectifying the PAM signal. Additional circuitry is necessary to accommodate the case where two adjacent channels' pulses have equal amplitudes and no switching transient occurs. In systems with a small number of channels, it is possible to synchronize each frame and use a sufficiently small desampling aperture that no error occurs. This synchronization method has proved adequate for a system with eight information channels and is the one chosen for the demodulator system.

4.3 Channel, Sync and AGC Pulse Levels

Sync and AGC pulses are included in the composite PAM waveform in addition to those from the signal channels. The sync pulse is recognized by the demodulating unit and used to synchronize the demultiplexing sampler with the multiplexing sampler in the transmitter. This pulse is made unique in the PAM wave train by setting its amplitude higher than any channel's pulse. Twenty percent of extra height was determined adequate by experiment.

The AGC pulse may be any fixed level in the composite PAM. A zero-amplitude value was chosen as this can be generated without the use of additional circuitry. Absence of multiplexing-gate output during one time slot gives a zero-level output.

4.4 Video Bandwidth

Finite video bandwidths give crosstalk between channels. The level of acceptable crosstalk (interfering signal in the output of one channel due to signals in other channels) determines the video bandwidth or frequency response for the system. Figure 4.1 gives a portion of a composite PAM waveform with a 100 percent duty cycle, along with a narrow desampling aperture. The rounded sides of the pulses are the result of a limited video bandwidth. Crosstalk occurs between adjacent channels when the transient, due to the switching between channels, does not completely die out before desampling occurs.

Assuming a transient response governed by a single time constant, a crosstalk of 0.5 percent requires a transient decay of 5.3 time-constants. ($e^{-t/\tau} = .005$ for $t/\tau = 5.3$, where t=time and τ =time constant). The rise time T_R is related to the time constant by

$$T_R = 2.2\tau \tag{4.1}$$

and the video bandwidth, B_v, is related to the rise time by

$$B_v = 0.35\ (1/T_R) \tag{4.2}$$

This result is a good approximation for higher order systems also[7]. For a sample window aperture, beginning at 50 percent of the sample period, the transient due to switching must have decayed to less than 0.5 percent in one-half of the sample period. This requires a rise-time of one-fifth of the sample period. Relating the required video bandwidth, B_v, to the sample period, T_s, gives

$$B_v = \frac{1.7}{T_s} \tag{4.3}$$

This requirement could be lowered by desampling later in the pulse period. The value in Equation (4.3) will be taken as a lower bound. Any wider bandwidths will result in better crosstalk rejection.

The low frequency response must also be considered. The time constants governing this must be long enough that none of the DC level is lost during the frame (One frame consists of a set of sample pulses, one from each channel plus a sync pulse and an AGC pulse). If this condition is not met, low frequency crosstalk occurs. For a crosstalk under 0.5 percent, the low-frequency time constant must be 200 times the length of the frame, T_F. Assuming that a single time constant governs this response requires a low frequency cutoff, f_1, of

$$f_1 = 8.0\ 10^{-4}(1/T_F) \quad . \tag{4.4}$$

4.5 RF-Carrier Bandwidth

The choice of RF-carrier bandwidth is dependent on available receiving equipment. Noise considerations require the use of the smallest bandwidth as noise voltage is proportional to the square root of the bandwidth.

Standard FM-stereo tuners have video bandwidth of about 55 kHz with IF amplifier bandwidths of about 200 kHz. Telemetry receivers have a wide range of bandwidths available. The receiver used in the development of this system was an Astro Communications Laboratories model TR-104. This receiver offers plug-in IF strips with bandwidths from 10 kHz to 3000 kHz. For FM, the 500 kHz IF strip offers 150 kHz video bandwidth and the 1000 kHz IF unit about 400 kHz video bandwidth.

The choice of RF bandwidth depends on the video bandwidth requirements of the PAM waveform and on the deviation ratio desired. Good noise-performance requires high values of deviation ratio and the efficient use of bandwidth requires low values. Therefore, the choice of RF bandwidth is a compromise including considerations of available receiving equipment, desired deviation ratio and video bandwidth, and RF-link noise considerations. This choice must be made simultaneously with the selection of deviation ratio for a predetermined video bandwidth.

4.6 Peak RF-Carrier Deviation and Deviation Ratio

The amplitude of the frequency deviation of the carrier is the peak carrier-deviation, f_D. This is related to the maximum video-bandwidth frequency, f_m, modulating the carrier, by the deviation ratio, D, in the relationship

$$D = f_D/f_m \tag{4.5}$$

The RF bandwidth necessary to preserve all the FM spectra above one percent is given in Figure 4.2 as a function of the deviation ratio. Figure 4.3 makes an alternate presentation of this curve, using units of modulating frequency, f_m, and plotting required bandwidth as a function of deviation ratio. For a system with a two percent accuracy, it has been experimentally found that these bandwidths can be reduced by about a factor of two[4].

The study resulting in the system design given in Table 4.1 shows that a deviation ratio of about two is optimum for a two percent accuracy system. As RF-bandwidth requirements become a limiting factor, lower values of deviation ratio become necessary to obtain the required video-bandwidth from a fixed RF-bandwidth.

4.7 Filters and Bandwidth-Limiting

It is necessary to use filtering in the signal channels to eliminate sampling errors. The frequency spectrum presented to the sampling gate must not possess components above one-half of the sampling frequency. With practical filtering a sampling ratio of four is used requiring a filter cut-off at one-fourth of the sampling frequency. This filtering will not be necessary in cases where the signals measured lack frequency components high enough to cause errors.

Video-bandwidth limiting by filters prior to the RF stage is necessary only when two or more transmitters will be operating at frequencies which are very close. In this instance the premodulation filtering eliminates side frequencies which would otherwise spill over into adjacent RF channels, causing interference between RF systems.

TABLE 4.1

DESIGN OUTLINE FOR TWO PERCENT ACCURACY

PAM-FM[4]

Parameters	Calculated	Measured
Transmitted SNR, $s/k_1 f_s^{1/2}$	14	14.7
RF Bandwidth	4 to $5 f_s$	4 to $5 f_s$
Video Bandwidth	$1.2 f_s$	f_s
Peak Frequency-Deviation	$2.3 f_s$	$1.75 f_s$
Duty-Cycle - Receive	0.7	0.6
Duty-Cycle - Transmit	1.0	1.0
Sampling Ratio	4 samples-cycle	------
Premodulation Filter	-----	$2.5 f_s$

f_s = sampling frequency, pulses per second

$\frac{1}{f_s}$ = period of each sample

k_1 = RMS noise per unit bandwidth

S = RMS carrier-voltage

SNR = Signal-to-noise ratio

Error Distribution

1. Fluctuation error of 1.4% - due to fluctuation noise at receiver input.

2. Crosstalk error of 0.6% - due to pulse spillover from previous channel.

3. Distortion error of 0.3% - due to nonlinearities, mainly limited receiver-bandwidth.

4. Sampling error of 1.2% - due to desampling at system output.

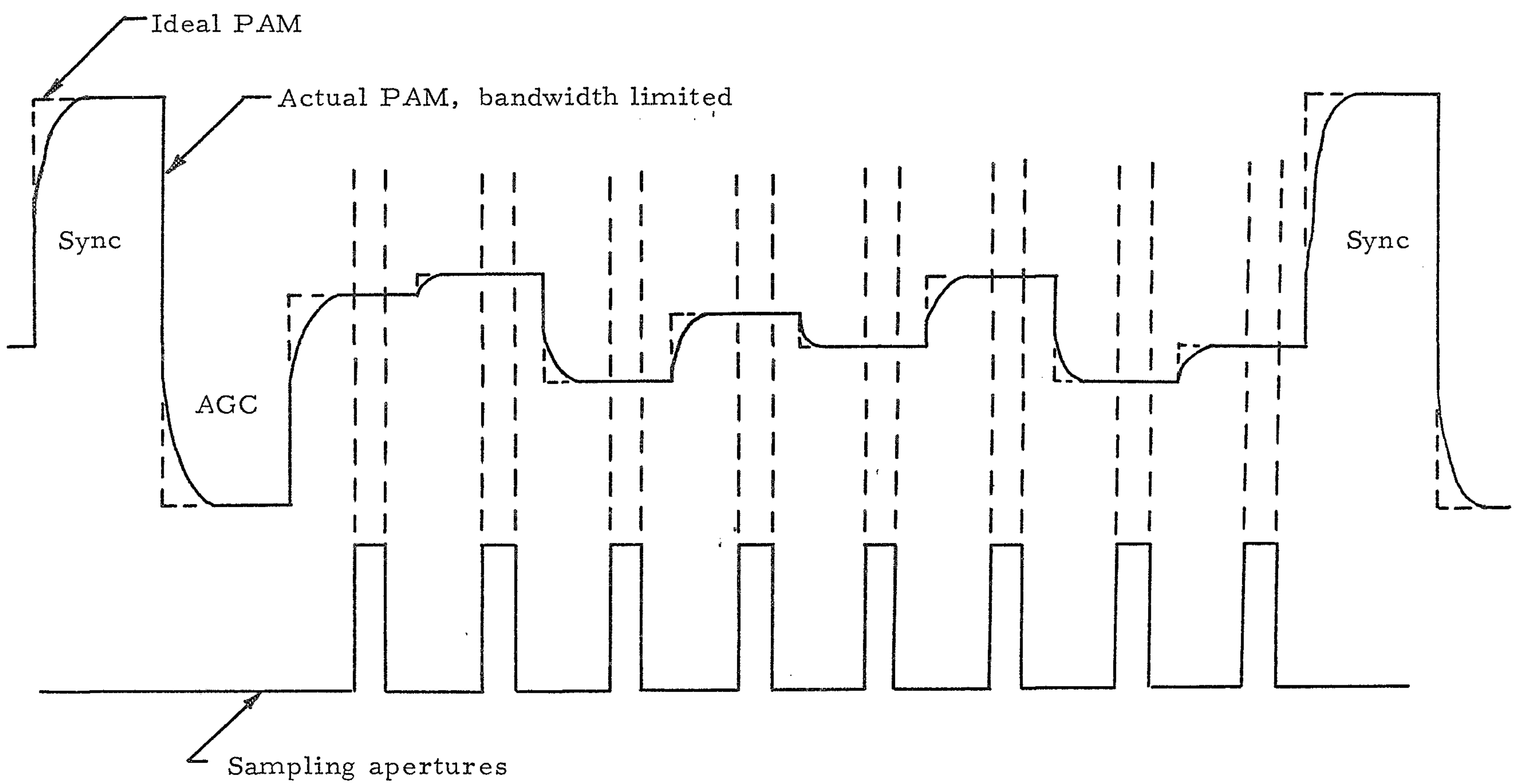


Figure 4.1. Composite-PAM waveform, showing rounding caused by bandwidth-limiting, and sampling aperture for demultiplexing gates.

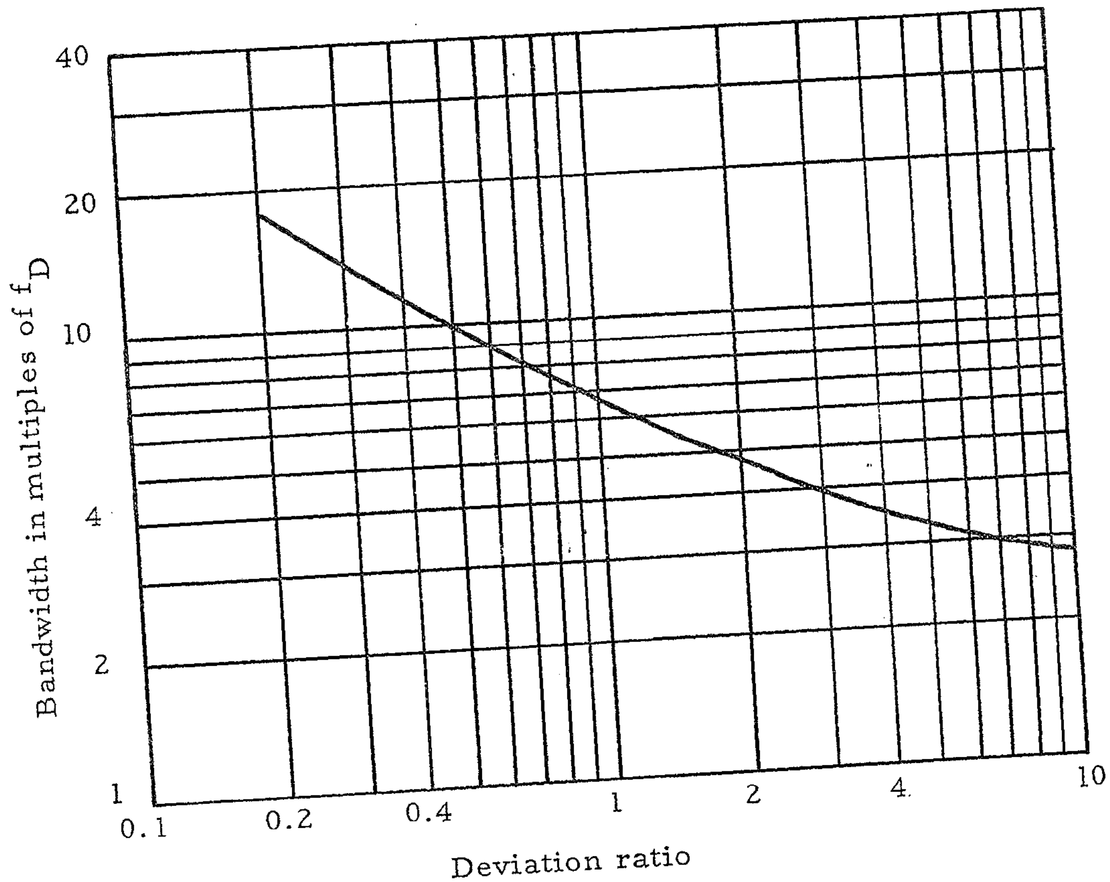


Figure 4.2. Bandwidth, to preserve sidebands above one percent, vs. deviation ratio[8].

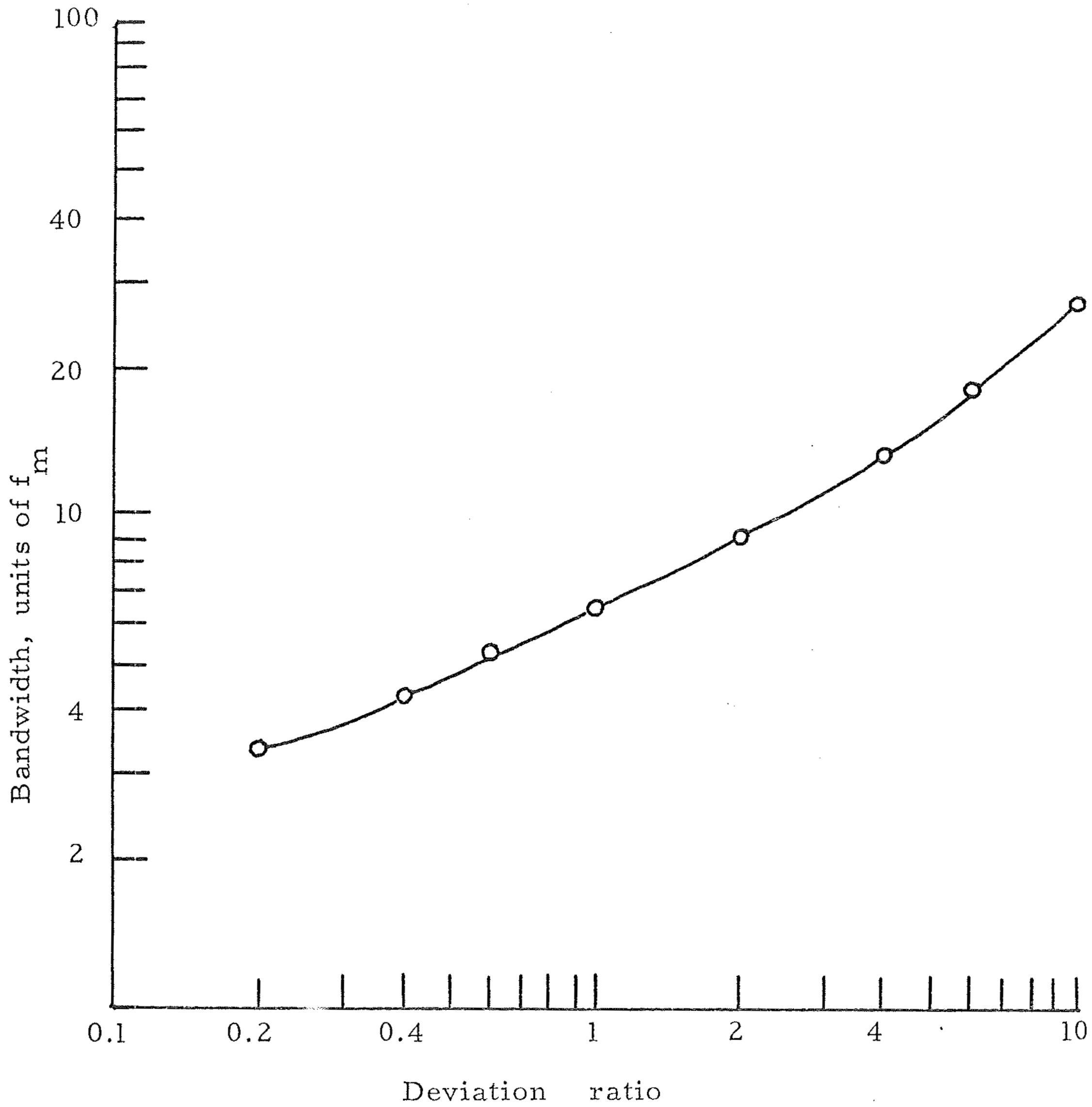


Figure 4.3 RF bandwidth, in units of modulating frequency, f_m, versus deviation ratio

SECTION V

CIRCUIT DESIGN TECHNIQUES

Circuit designs are limited by the requirements set forth in Section II. Available components, circuit simplicity, and current drain from the power supply are the major restrictions. Although the physiological ambient makes stringent requirements on the circuitry, it offers two very distinct and useful advantages - good shielding from external noise, and excellent temperature regulation.

5.1 Techniques for minimizing power requirements

Two techniques are available which minimize power drain in a given circuit. One is to design the circuits to draw very little power. The other is to use circuitry which draws appreciable power in operation but zero power when its operation is not absolutely required.

Micropower circuitry requires the use of very large resistances, in the order of megohms. These circuits are slow in their operation due to the large values of resistance. These resistance values are presently not feasible in hybrid integrated circuit form. For these reasons this technique is not applicable to this system at the present time.

Circuitry is available which requires no standby power. An example of this, given in Figure 5.1, is a complementary, PNP-NPN, bistable circuit. No current is drawn from the power supply when the transistors are turned off. The other state, with both transistors conducting, draws current, limited by the load resistance in the circuit.

Standby power can be eliminated in other circuits which normally consume power 100 percent of the time. This is accomplished by switching off the power to these circuits when their operation is not required. Power requirements are reduced by a factor equal to the duty cycle necessary for that circuit's operation. Alternately, this technique allows n circuits with equal power-requirements and a duty cycle of 1/n to function for the average, steady-state, power drain of one of the circuits.

5.1.1 Requirements on power-supply-gated circuits

Applying power-supply gating to linear circuits, as amplifiers and signal conditioners, makes certain demands on the circuits' design and components. In order for the circuits to be useful in this mode of operation, they must reach "steady-state" conditions (equivalent to non-gated operation) in a small portion of their use period or "on" period.

The turn-on time (time to reach steady-state "on" conditions, starting from the "off" conditions), will be equivalent to a delay in the beginning of the sample pulse as seen by the demodulator. The turn-on of the sample pulse will delay the end of the switching transient presented to the video bandwidth and consequently delay the end of the transient in the demodulator. To accommodate this, the video bandwidth must either be increased or the desampling time moved farther from the beginning of the channel time.

A circuit-turn-on time of less than five percent of the total "on" period will be negligible. The sample aperture beginning at 50 percent of the channel-time, as used in obtaining Equation 4. 3 can be moved to the 60 percent point of the channel time with no other change provided the aperture is sufficiently short.

5. 1. 2 Component limitations in gated-power supply circuits

The circuitry must switch from "off" to its steady-state operation in less than five percent of the sample period. This requires the individual transistors to be capable of switching in a small fraction of this time. The resistances in the circuit must be small so that the time constants, associated with them during turn-on, will be small. Capacitances should be avoided if possible. They may be included if special precautions are taken. It can be noted that these requirements are nearly identical with the requirements and achievements of integrated circuitry designed for high-speed or wide-bandwidth operation.

Capacitors may be used provided that a charge can be established and maintained on them equivalent to that reached in a non-gated-power operation. This requires that charge be conserved in the "off" portion of the power duty-cycle. The circuitry must be designed so that each capacitor sees an infinite impedance during this period. Electrically opening all connections to the circuit involved except to the capacitor, or opening one lead of the capacitor, will accomplish this.

5.1.3 Additional advantages and disadvantages of power-supply-gating

The turn-on time for the circuit when the entire transmitter is initially turned on will be longer than for non-gated-power operation. Charging current for the capacitors is provided only during the "on" period so that the initial turn-on time is multiplied by the reciprocal of the circuit's duty-cycle.

The long turn-on characteristic is equivalent to multiplying the time constant associated with a capacitor by the reciprocal of the duty-cycle. This offers attractive possibilities in circuit designs which are limited to capacitors with small physical size. A duty cycle of one-tenth will multiply the apparent value of a capacitor by ten times. In a capacitor-coupled, high-gain-pre-amplifier this can result in better low-frequency response with no increase in volume.

Duty-cycle multiplication of the time constant can be used to boost the input impedance of an amplifier. Figure 5.2 gives a circuit configuration for this. The capacitor is charged 100 percent of the time from the source. The amplifier withdraws charge only during its "on" period. The capacitor averages this drain over 100 percent of the time and sees an average drain equivalent to the peak drain multiplied by the duty cycle. Since this drain could be accomplished by an impedance equal to the actual input impedance divided by the duty cycle, the apparent input impedance is increased over the actual value.

The circuitry used for this impedance multiplication also performs as a low pass filter, serving two very important functions. It provides bandwidth limiting, prior to sampling, to reduce sampling errors. Also, it provides bandwidth limiting which allows cascading of high-gain amplifiers. This eliminates oscillations between the two amplifiers. Differential amplifiers provide isolation between power supply and signal lines at low frequencies, but as frequencies increase this isolation decreases. Oscillations usually occur at frequencies well above the signal frequencies, interfering with normal operation. The series low-pass filter reduces high-frequency gain to eliminate these oscillations.

Another, and very important, feature of power-supply gating, is in the area of crosstalk elimination. If all amplifiers are turned off unless their channel is being sampled, there will be no signals present in the transmitter of sufficient magnitude to generate any crosstalk.

5.2 Techniques and components

The components available may be summarized as:

a) Wide range of transistor parameters and units

b) Some monolithic integrated circuits available

c) Resistances up to 50 kΩ. Ratios available with high tolerances.

d) Very few capacitors; chips up to 500 pf.

e) No inductors

Transistors should be used in as many locations in the circuitry as possible. It is better to use a transistor than a passive device if at all possible. A wide selection of transistors is available in chip form and transistors are the cheapest component available in this form.

The use of power-supply gating allows the use of some monolithic integrated circuits. Low-duty-cycle operation will reduce their power requirements and make them compatible with the overall system power requirements.

Resistances up to 50 kΩ are available on a single chip. Ratios of resistances are obtainable from a single chip to a high degree of accuracy. The absolute resistance-values may vary by 50 percent from one chip to another but ratios are dependent almost entirely on geometry and one percent tolerances are easily obtained.

Capacitors and inductors should be avoided if possible. Capacitor chips are available up to about 500 pf. These values may be multiplied by emitter followers and by duty-cycle multiplication.

A major portion of the transmitter, which is discussed in Section VI, is a ring oscillator. This provides the control signals for power-supply-gating and for sampling and multiplexing. The memory units for the ring oscillator are PNP-NPN, bistable stages, similar to the circuit in Figure 5.1. These memory units were originally interconnected, using resistor-transistor logic.

A revised design uses capacitive coupling between stages to eliminate the need to select transistors with different switching speeds to accomplish the required time delays. Also, a single unit, an SCS (silicon controlled switch), is used to replace the complementary pair of transistors required to perform the memory function.

Resistances are used to limit the currents in the circuits. The circuits are designed so that the absolute values of most of the resistances can vary by +100 percent to -50 percent and not affect the operation of the circuit.

Noise immunity in the ring oscillator is provided by saturated logic and low impedance-levels. Also, in the "off" state the circuits have zero gain to any disturbances below a level sufficient to turn on a base-emitter current.

5.3 Advantages of physiologically-implanted ambient

A transmitter which has been implanted in a mammal resides in a conductive medium which has very good thermal-regulation properties. The conductive medium provides a degree of isolation from external electric and magnetic fields. Temperatures in a healthy animal will not vary more than about $2^{o}C$. Circuitry will remain quite stable over this limited temperature range. The circuit designs need accommodate room-temperature operation and implant-temperature operation, a range of about $20^{o}C$.

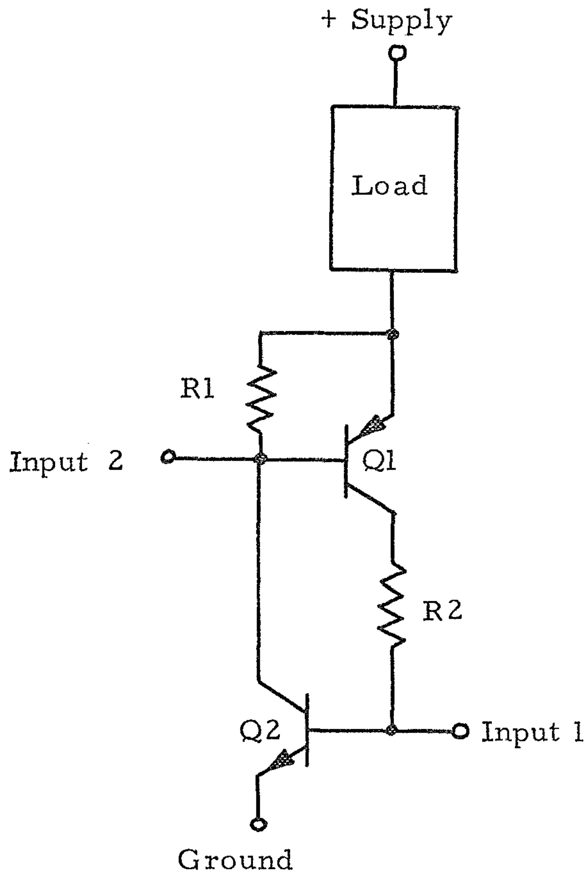


Figure 5.1 PNP-NPN complementary bistable circuit.

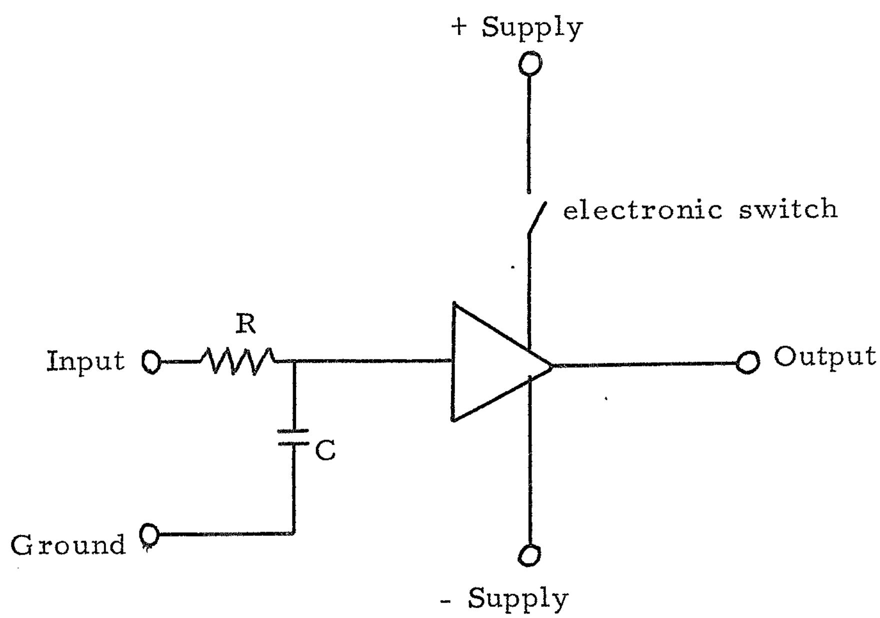


Figure 5.2 Input impedance multiplication by capacitor in gated-power-supply operation.

SECTION VI

TRANSMITTER DESIGN

A transmitter for a physiological implantable, PAM-FM, telemetry system must contain the following sections:

a) Sampling and multiplexing control source

b) Signal conditioners and amplifiers

c) Sync and AGC generators

d) Multiplexing gates

e) RF circuitry

f) Power supply

g) Remote switch

This section will discuss the features of the transmitter unit in functional block form. Details of the circuitry and its operation are also given and discussed in this section.

6.1 Block diagrams of the transmitter unit

The transmitter circuitry is shown in block-diagram form in Figure 6.1. The ring oscillator controls the sequential sampling of the channels, their multiplexing, and the gating of the power supplies to their signal conditioners and amplifiers.

The signal conditioners include the input transducers or pick-ups necessary to convert the physiological signals to electrical signals. The amplifiers raise the levels of the electrical signals to that sufficient for the multiplexing gates. The multiplexing gates connect the amplifiers' outputs, in sequence, to the common PAM-line. The composite PAM signal then modulates the RF oscillator. The oscillator radiates to the external receiver.

Power, to operate the transmitter, comes from the power pack. The remotely-operated switch allows the transmitter to be operated only when studies are being made, thus conserving power and allowing a finite-capacity power pack to be used intermittently over a long period of time.

The frame-sync generator provides a pulse, higher than any channel's pulse, in the composite-PAM waveform. This pulse is recognized in the demodulator and used to synchronize its demultiplexing gate with the multiplexing gate of the transmitter.

The AGC generator provides a fixed level in the composite PAM. This level in conjunction with the frame-sync pulse's level carries DC-level and gain-calibration information.

6.2 Ring oscillator

The ring oscillator consists of a ring counter, a clock or driver circuit, and an automatic-reset circuit. A block diagram of the ring oscillator is given in Figure 6.2.

The ring counter stages employ PNP-NPN, complementary, flip-flops as memory units. Resistor-transistor logic gates couple the stages together into a ring. There are no capacitors in any of this circuitry. The circuitry is given in Figures 6.3 and 6.4.

Each stage has an "and" gate for a trigger or clock signal input. The presence of an "on" state in one stage and a trigger signal causes the next stage to be turned on. An "or" gate in each stage is connected to the "sense" output of the following stage. This turns off a stage when the following stage turns on.

There is a short period during the switching action when the two stages involved are on simultaneously. This is the time required to turn off the first stage when the second is triggered on. This overlap is sensed by an "analog and" circuit and a threshold detector and is used to turn off the trigger pulse. In so doing, it triggers the monostable circuit which serves as the clock for the ring oscillator. The diagram for the "analog and" circuit is given in Figure 6.5.

Closed-loop operation insures that the trigger pulse is long enough to shift the ring counter. The pulse must not be too long. If it were, some succeeding stages could be triggered on with a single-trigger-pulse. Thus, it is necessary to achieve a balance between the switching speed of the ring oscillator stages and the trigger-generating loop.

This timing balance will be satisfied if the trigger pulse, shifting the ring oscillator from stage n to stage n + 1, is turned off before the "sense output" of stage n + 1 is turned on. If "sense output" comes on before the trigger pulse is turned off, stage n + 2 may be triggered on. The net result is a shift of two stages instead of one. Another possibility is for the trigger pulse to be turned off before sufficient time has elapsed to allow the bistable memory to switch. Using two relatively slow transistors in the circuit loop involved in turning on a stage and fast ones in the trigger-cut-off loop, can achieve this balance.

The preceding discussion covers the normal, steady-state operation of the ring oscillator. It is necessary to provide a means of starting this kind of circuitry, and for restarting it if a malfunction occurs. This ring oscillator inherently will not function if there is no stage on or if two or more non-adjacent stages are on. When either of these cases occur, the trigger generator will be held either on or off and will not switch. Differentiating the trigger pulse provides a signal which indicates proper operation. This signal is used to inhibit the operation of a monostable circuit similar to the clock generator but with a period five to ten times longer. Each differentiated trigger pulse restarts this longer delay. The absence of pulses allows its output to turn on at the end of the delay.

An automatic-reset feature is obtained from this circuit. It only turns on when the ring oscillator fails to operate. An "or" gate on each individual ring counter stage is connected to the output of the automatic-reset stage. One stage is connected so that the reset gate turns it on. The other stages are turned off. Once this condition is obtained, the normal operation of the ring oscillator resumes.

The diagrams for all of the circuits used in the ring oscillator are given in Figures 6.3 through 6.7.

6.3 Revised Ring Oscillator

A block diagram of the revised ring oscillator is given in Figure 6.8. Figure 6.9 is the diagram for the new ring oscillator driver circuit. The SCS (3N84) is operated as a programmable unijunction transistor and the timing capacitor is charged from a constant current source formed by Q_1, Q_2 and the associated resistors. Q_3 is a capacitance multiplier to increase the amount of time required for the SCS to discharge the capacitor. Q_4 and Q_5 form an isolation and squaring amplifier for the pulse output from the anode gate of the SCS. External connections are provided in parallel with the emitter resistor of Q_2 to allow adjustment of the charging current into the capacitor, which in turn controls the frame width of each channel.

Figure 6.10 gives the diagram of a flat pack containing two ring oscillator stages. Q_1 and its associated biasing network are a low value current source. This current source keeps leakage current through the capacitor from holding the SCS on and has a high enough effective resistance to allow almost all of the pulse current through the capacitor to go into the SCS. Q_3 is a clamping diode to prevent the cathode gate of the SCS from being forced too far negative by the leading edge of the preceding stage pulse. Q_2 is the switch which applies power to the appropriate signal input amplifier. The base-emitter voltage drop of Q_2 also provides biasing for Q_1.

The first ring oscillator stage is shown in Figure 6.11. The automatic restart circuit initially turns the SCS on, and turns it on each time the trigger line goes above the saturation voltage of the SCS. After the pulse is shifted out of the last ring oscillator stage, the trigger line tries to rise to the supply voltage. The automatic restart circuit then switches the first stage on again to start a new cycle. Since this circuit requires only two chips, a tapped resistor, and an NPN transistor, and the input coupling capacitor is no longer necessary, the automatic restart circuit is included in the flat pack with the first two ring oscillator stages.

Operation of the ring oscillator takes place in the following manner. When the circuitry is initially turned on, one or more of the SCS's may be turned on by this transient. If so, these "on" states are shifted out of counter by successive trigger pulses.

Once the ring oscillator stages are cleared, normal operation can take place, starting with the automatic restart circuit turning on the first ring oscillator stage. Figures 6.12 and 6.13 show waveforms obtained from a properly operating ring oscillator. Figure 6.12 gives an overall idea of the operating cycle and the sequence of events. Figure 6.13 gives more detail as to some of the transients which occur.

It has been found that the value of the coupling capacitor between ring oscillator stages is non-critical as long as its value exceeds 150 pf.

6.4 Signal conditioners and amplifiers

The design of signal conditioners and amplifiers is dependent on the characteristics of the signals to be telemetered. The signal conditioner for a particular channel includes the transducer necessary to sense the desired signal and the circuitry required for the transducer. An amplifier is used to raise the level of the electrical output from the signal conditioner to the value required by the multiplexing gates.

Circuitry has been designed for two types of inputs. Resistance-bridge input circuitry for strain gauges, thermistors, etc., has been developed. Amplifiers have also been developed for direct-electrical signals as EKG, EEG, and EMG.

Figure 6.14 gives the basic circuitry for a resistance-bridge channel. The differential amplifier is a Fairchild μA 702, available in chip form. This amplifier works well in gated-power-supply operation and with 6-volts-total power supply. A stable gain of 100 can be achieved without any capacitors. The power for the amplifier and for the input bridge is controlled by a ring oscillator output.

The modified strain gauge amplifier is shown in Figure 6.15. This circuit has reduced power drain, and the output was modified to accommodate switching the negative supply instead of the positive supply voltage. Q_1 and Q_2 are a differential amplifier with a usable voltage gain of about 20. Q_3 and Q_4 form a constant current source for the differential amplifier, the current being determined by the value of R_7. R_5 and R_6 form half of the bridge circuit. The other two arms of the bridge circuit are formed by either a strain gauge or some other variable resistance sensor. Q_5 is a pass transistor which cuts off the output as soon as the next ring oscillator stage is turned on. Q_6 provides a level shift and additional isolation between the differential amplifier and the output to the modulator circuit.

Q_7 is the output transistor which drives a common line to the modulator sync generator circuit. The gain of individual channels can be adjusted by varying R10. The overall gain of all the channels is adjusted by varying the common collector load resistor which is in the modulator circuit.

Basic circuitry for a high-gain amplifier to handle direct electrical signals is given in Figure 6.16. The first differential amplifier is low power and operates full time. It is designed to be placed at the electrode location, when used, eliminating noise in long input leads by sending an amplified signal at lower impedance back to the main transmitter package. The second differential amplifier, in the cascaded pair, is operated with gated power supply.

The frequency response of the channel is shaped by capacitors C_1, C_2, and C_3. C_1 limits the low-frequency response and blocks the DC level at the output of the first stage from the input to the second. This eliminates effects of DC unbalances in the electrodes and in the first stage which would saturate the second stage. Capacitors C_2 and C_3 provide low pass filtering ahead of the sampling gate which is connected to the output of the second stage to prevent sampling errors. Capacitor C_3 eliminates high frequency oscillations between the two stages as well as multiplying the input impedance of the second stage.

Other input transducers could be accommodated within the framework encompassed by these two basic channels. Given a transducer, the problem would be in matching it to one of these amplifiers.

6.5 Frame-sync and AGC generators

The frame-sync generator is a multiplexing gate which looks at a fixed level. The output pulse from this generator is set about 20 percent higher than the maximum height pulse of any channel. The AGC pulse is a zero-level pulse. This was chosen since it can be generated by the absence of any circuitry. The schematic diagram for the frame-sync generator is given in Figure 6.17.

In order to minimize the number of PNP transistors necessary in the revised transmitter, and yet keep the modulation referenced to the negative supply line, a modulator and sync generator circuit was designed to accept the outputs of the modified strain gauge amplifiers and apply them along with a sync and AGC pulse to the r.f. oscillator unit. The schematic diagram for this circuit is shown in Figure 6.18. Q_3 is the modulator. The collector load resistor is set when the entire transmitter is assembled to set the gain of the modulator and thus the deviation of the RF oscillator. Q_2 provides the sync level to the modulator when the first ring oscillator stage output grounds the emitter resistor. Provision is also made to adjust the height of the sync plus at final assembly time.

R_1 is the load resistor for the second ring oscillator stage and Q_1 is a clamping diode to insure that the sync level terminates when the second ring oscillator stage turns on.

6.6 Multiplexing gates

The basic unit of the multiplexing gate is given in Figure 6.19. This circuit supplies a current to the PAM line proportional to the channel's output. The overlap gate, Q_2, turns off the multiplexing gate when the following stage comes on. This minimizes overlap in the PAM waveform due to ring oscillator overlap.

This gate offers excellent isolation between the PAM line and the channel's output when it is turned off. The PAM line has signals under 300 mv. maximum and these levels will not cause the collector-base junction of the silicon transistor to conduct. The amplifier's output similarly sees an open circuit.

The output of the multiplexing gate is limited so that pulses big enough to interfere with the sync pulse are not generated. Resistor R_1 limits the current delivered from the gate due to saturation of the amplifier or due to a malfunction in the channel.

A higher-gain multiplexing gate possessing the same limiting and isolating features is given in Figure 6.20. When the differential amplifier saturates the gate becomes identical to the lower-gain one of Figure 6.19. This gate requires close balancing of the preceding amplifier but yields an appreciably higher gain.

6.7 RF Circuitry

The schematic diagram of the RF circuitry is given in Figure 6.21. The RF stage is a varicap-modulated Colpitts oscillator. Q_1 is the oscillating transistor. C_4 and C_5 provide feedback. C_1 and C_6 are bypass capacitors. C_2 and C_3 are the modulating capacitors. The RF coil is connected between points D and E and is external to the integrated-circuit package.

The diodes in the circuit are base-emitter junctions of transistors. D_3, D_4, R_3 and R_4 combine to bias Q_1 with a constant current to minimize effects from power-supply variations. R_2 provides isolation between the RF voltages and the PAM line connected to the point B. C_1 also provides low-pass filtering prior to modulation of the RF. R_1 develops the modulation voltage for the varicaps from the current supplied by the PAM line. R_1 is adjusted to vary the modulating sensitivity of the stage.

Diodes D_1 and D_2 provide about one volt of power-supply reduction since the total power supply is equal to the voltage-breakdown of the chip capacitors used. With the revised circuitry the voltage reduction is unnecessary, and these two diodes have been eliminated from the RF flat pack. This statement will be explained in the next section.

6.8 Power Supply

A power pack containing five mercury cells was chosen for the initial transmitters to be implanted. This gave a tapped power supply of plus 3.9 volts and minus 2.6 volts and was the minimum number of cells that would allow the Fairchild μA702's to function. Nickel-cadmium rechargeable batteries, with a voltage of 1.25 volts per cell, are close to the potentials offered by mercury cells. These two types could be interchanged with no other circuitry changes. This allows for a rechargeable power pack when remote-charging circuit is developed. Nickel-cadmium cells were used in testing and developing the system and mercury cells in the implanted transmitters.

The power pack for the revised circuitry was chosen to be four mercury cells yielding an operating potential of 5.2 volts. These circuits will also work with five cells. A minimum of about 4.5 volts is required for reliable operation of the ring oscillator.

6.9 Remote switch

Remote switching of the transmitter's power is effected by a bistable electronic switch triggered with a magnetic reed-switch. An external magnet is used to close the reed switch and cause the electronic switch to change state. Studies were also made using a small bias magnet in conjunction with a reed switch, giving a polarized and latching reed switch. This method was rejected since a strong external field can demagnetize the bias magnet, killing the operation of the circuit. The diagram of the remote-switch circuitry is given in Figure 6.22.

Since the revised circuitry does not require switching of both the plus and minus power supply leads, a simple latching magnetic reed switch is used to turn the circuitry on and off.

6.10 Transmitter Subcommutation

Circuitry has been breadboarded which would allow the subcommutation of one of the information channels in the transmitter. Figure 6.23 shows the circuitry necessary to modify one of the existing strain gauge amplifiers for use in a subcommutating system. The actual power for operation is obtained from the lower speed driven ring oscillator stage. The output of the primary ring oscillator stage which is to be subcommutated is used to key the other input so that power is applied only during the frame when it is required. A large resistor is in series with the keying signal to keep this keying signal from applying sufficient power to operate the subcommutated amplifiers. This series resistor also helps prevent loading of the output of that ring oscillator stage.

Figure 6. 24 shows the circuitry used for the ring oscillator driver for the subcommutating ring oscillator. This circuit just differentiates and inverts the output of one of the main ring oscillator stages. This spike is then applied to the shift line of the subcommutating ring oscillator stages, and they operate similarly to the main ring oscillator.

This circuitry was developed for the original ring oscillator. Similar circuitry could be developed for the revised ring oscillator with a few modifications.

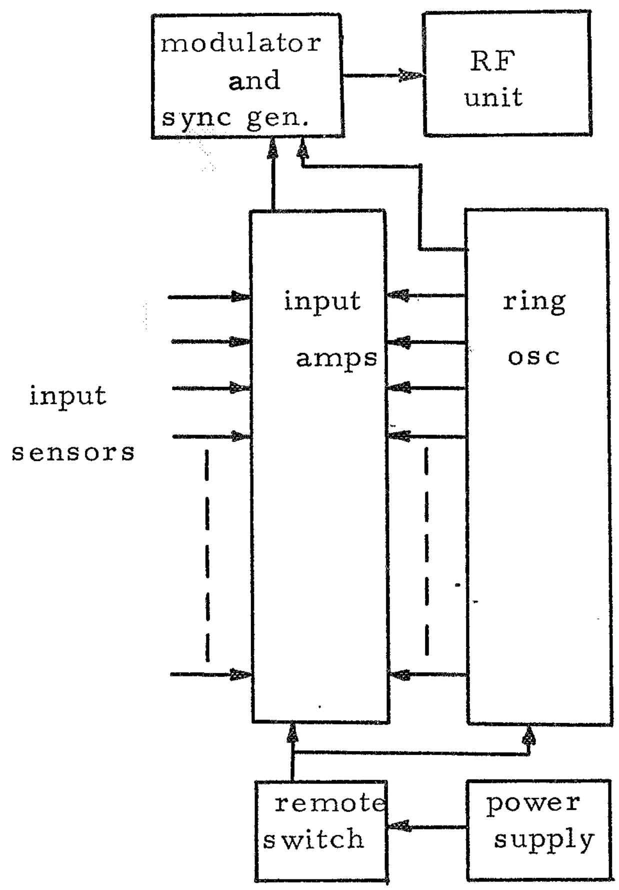


Figure 6.1 Block Diagram of a Transmitter

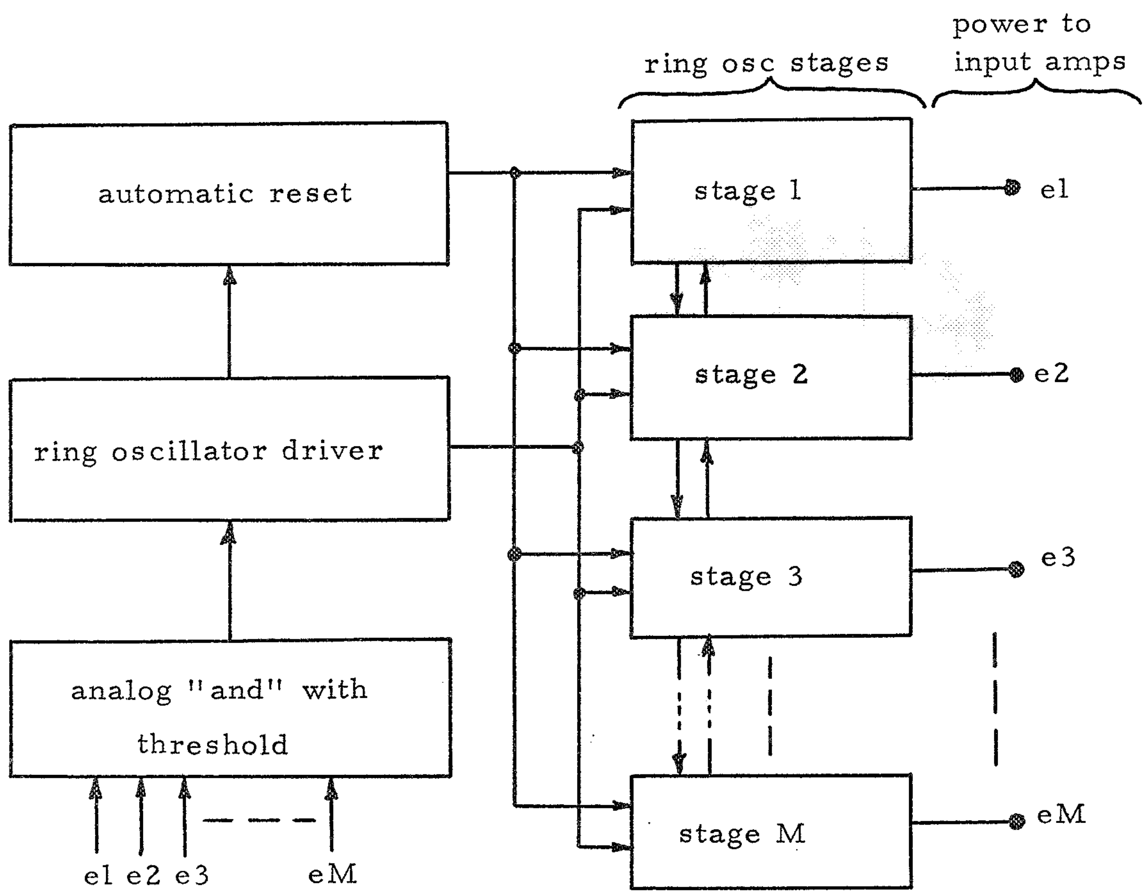


Figure 6.2 Ring Oscillator Block Diagram.

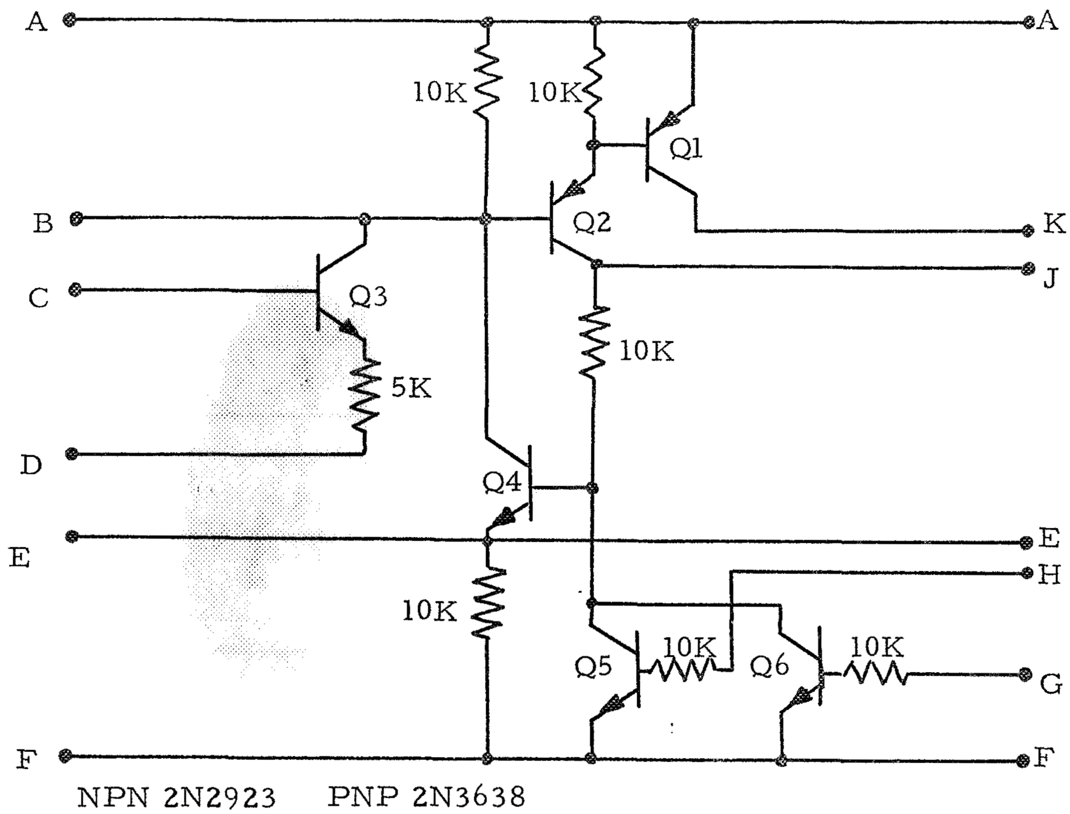


Figure 6.3 Ring-oscillator stage with reset-to-off input

A) +4 volts
B) Test point
C) To sense output, previous stage
D) To trigger line
E) Sense output
F) O volts
G) To reset line
H) To sense output, following stage
J) Test point
K) Output

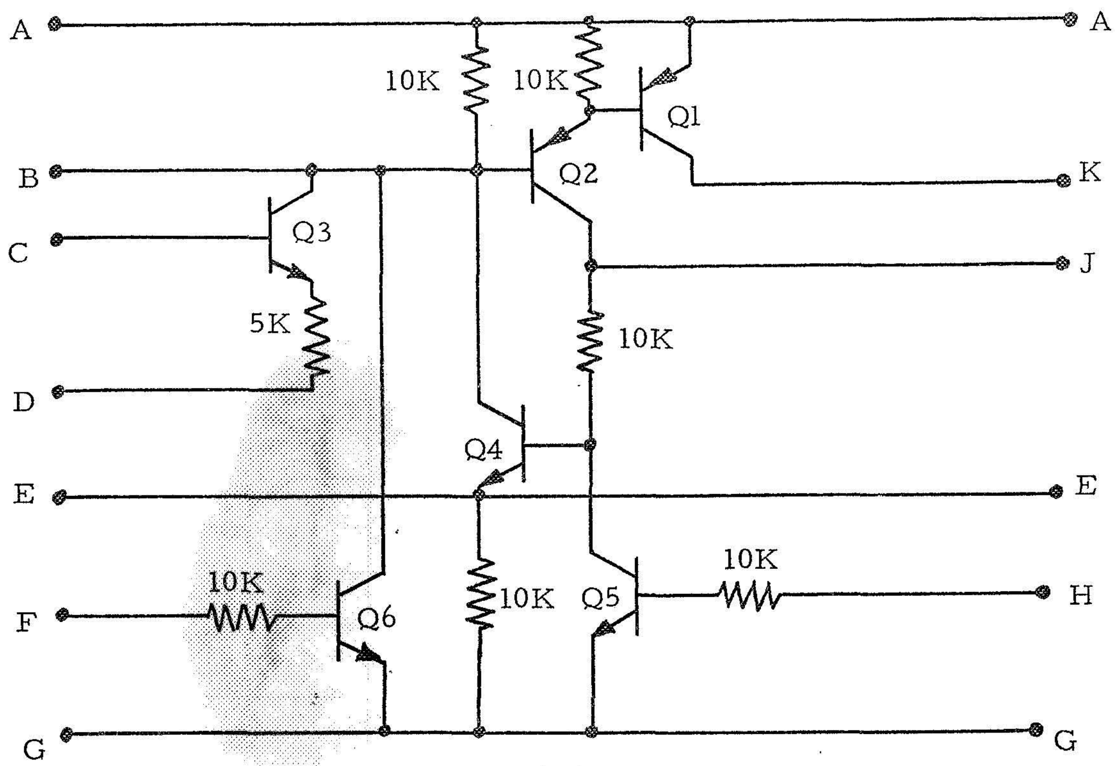
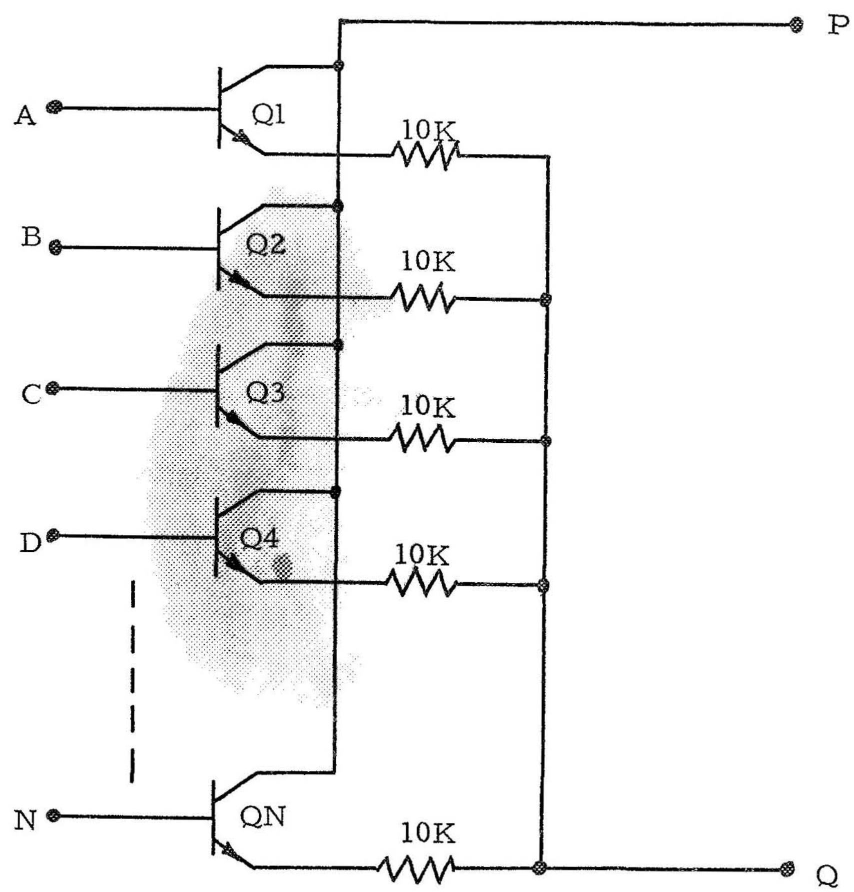


Figure 6.4 Ring-oscillator stage with reset-to-on input

A) + 4 volts
B) Test point
C) To sense output, previous stage
D) To trigger line
E) Sense output
F) To reset line
G) O volts
H) To sense output, following stage
J) Test point
K) Output



all transistors 2N2923

Figure 6.5 "Analog-and" circuit for ring oscillator

A) to N) Inputs, from ring-oscillator-
stage outputs
P) Output
Q) + 4 volts

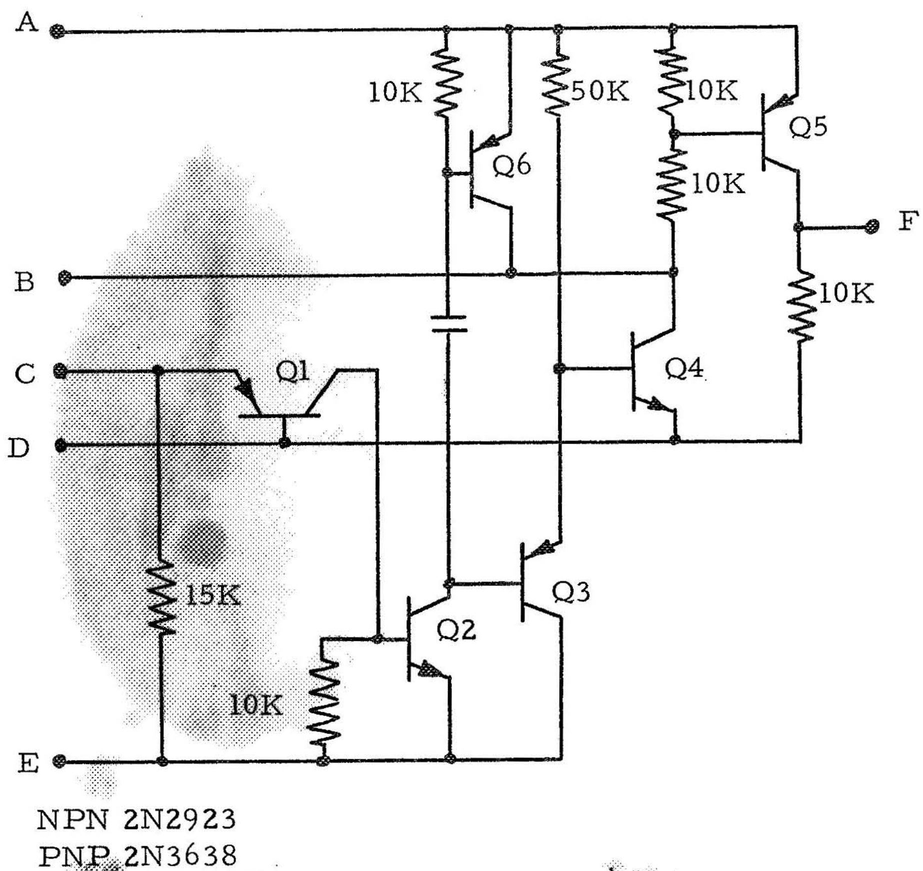
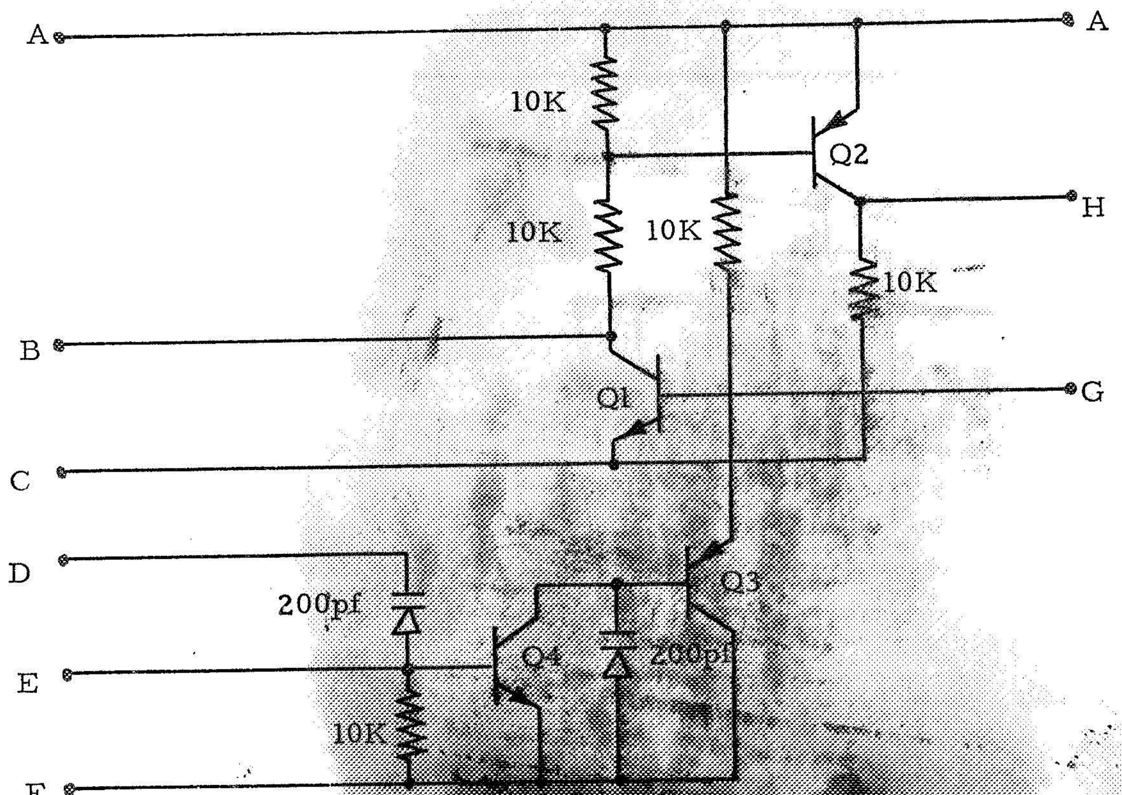


Figure 6. 6 Ring-oscillator driver

A) + 4 volts
B) Trigger-line output
C) Input
D) O volts
E) - 2. 5 volts
F) Output to automatic-reset circuit



NPN 2N292 [illegible]
PNP 2N3638 [illegible]

Figure 6.7 A [illegible]

A) +4 [illegible]
B) Test p [illegible]
C) [illegible]
D) Inp [illegible]
E) Input, [illegible]
F) - 2.5 [illegible]
G) Test p [illegible]
H) O [illegible]

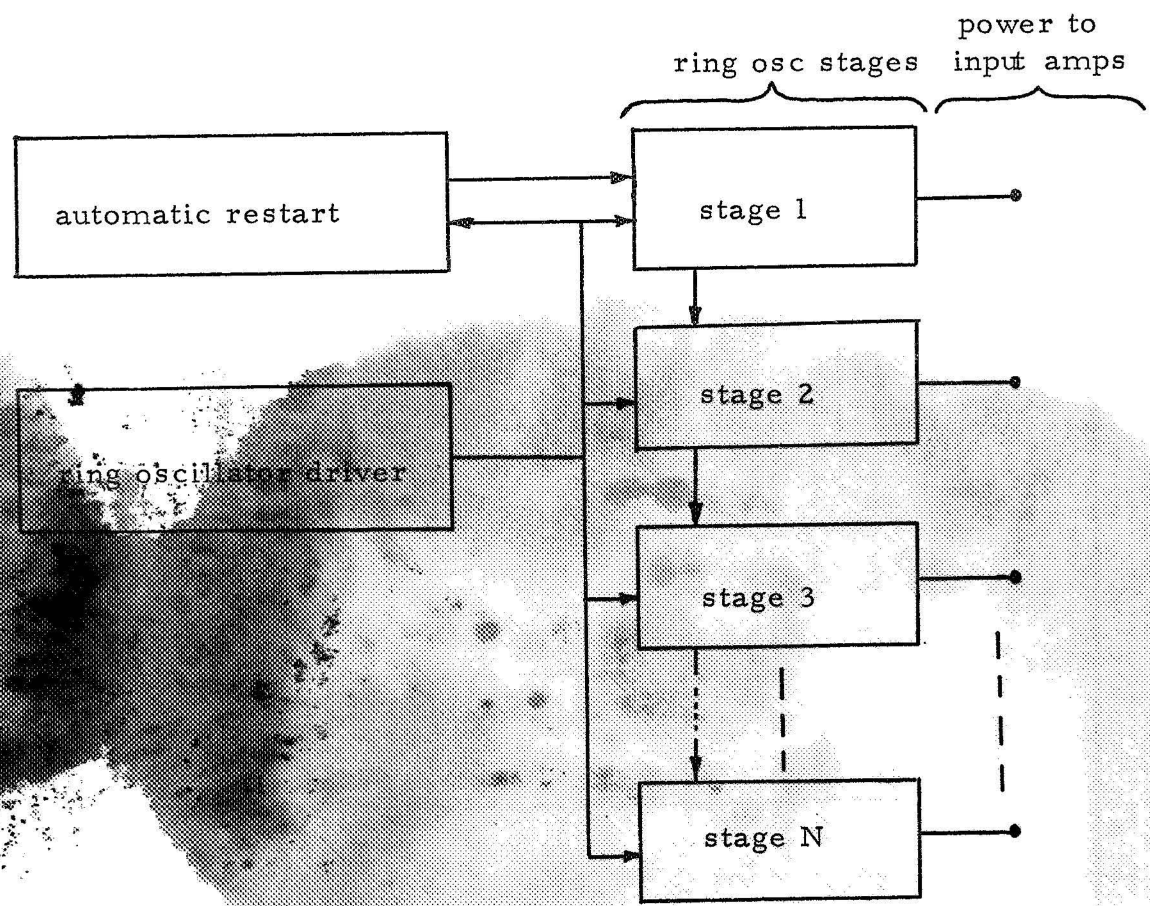
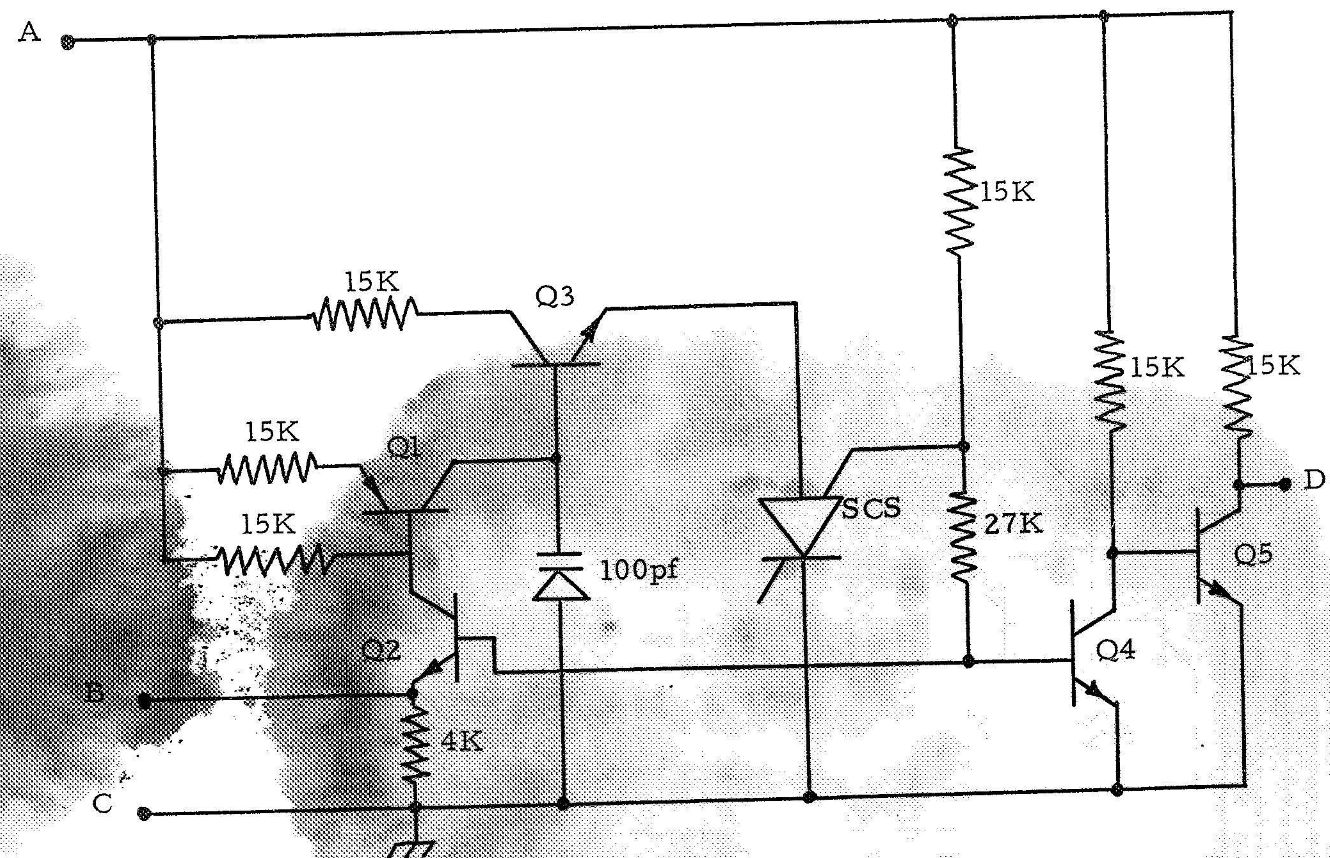


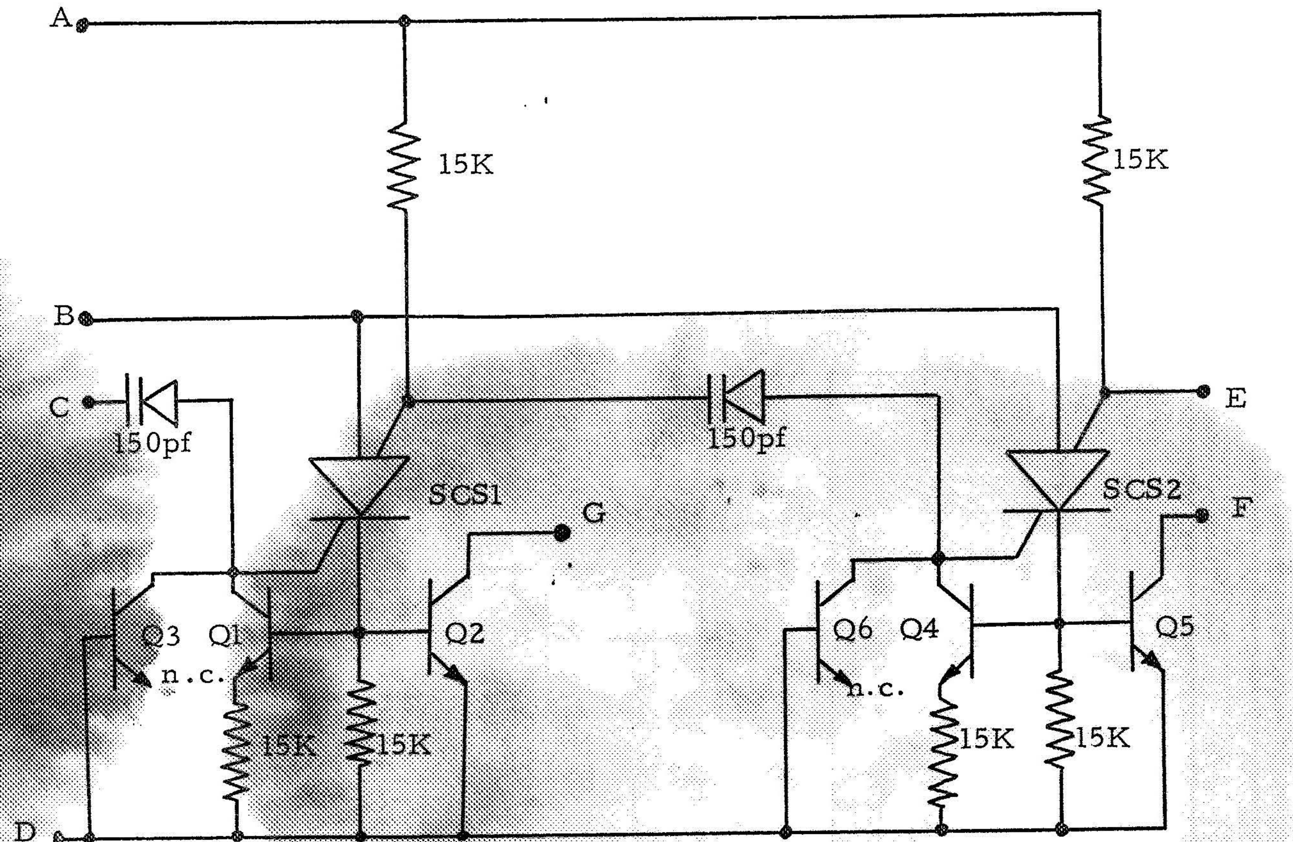
Figure 6.8 Block Diagram of revised ring oscillator



A) +5 volt supply
B) timing adj
C) common
D) trigger line output

PNP 2N3638
NPN 2N3663
SCS 3N84

Figure 6.9 Ring oscillator driver circuit

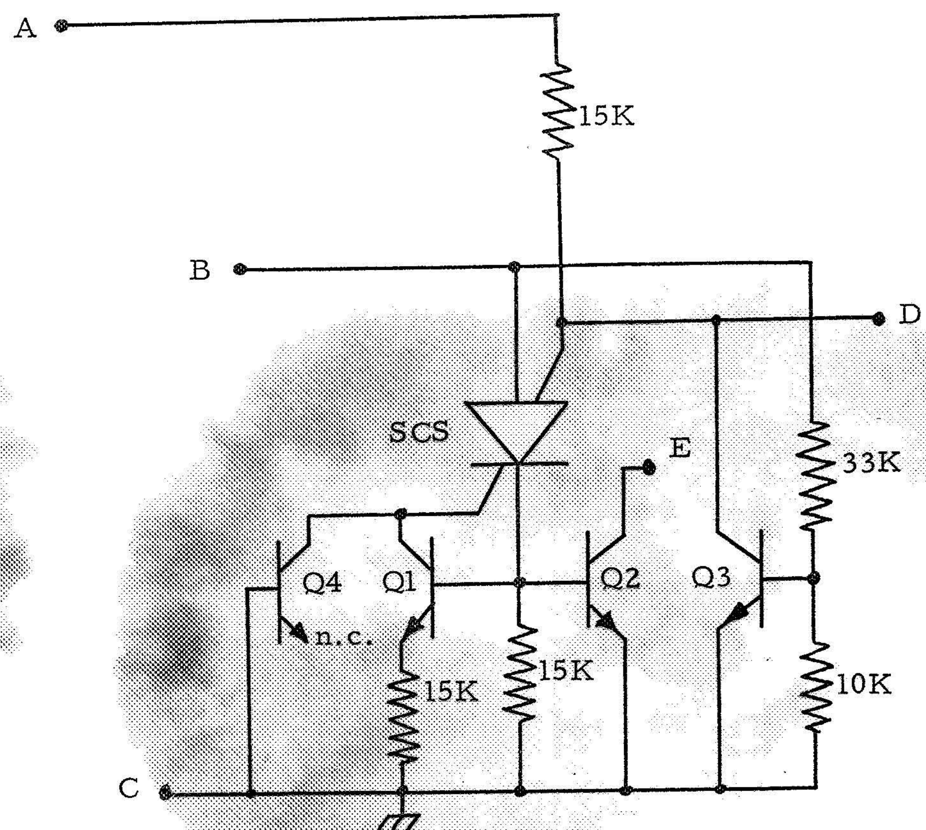


A) +5 volt supply
B) trigger line
C) from previous stage

D) common
E) output to next stage
F) output to input amp B

G) output to input amp A
all transistors 2N3663
SCS1 + SCS2 3N84

Figure 6.10 Two ring oscillator stage circuits



A) +5 volt supply
B) trigger line
C) common
D) output to next stage
E) output to input amp
all transistors 2N3663 SCS 3N84

Figure 6.11 First ring oscillator stage with automatic restart circuit

A

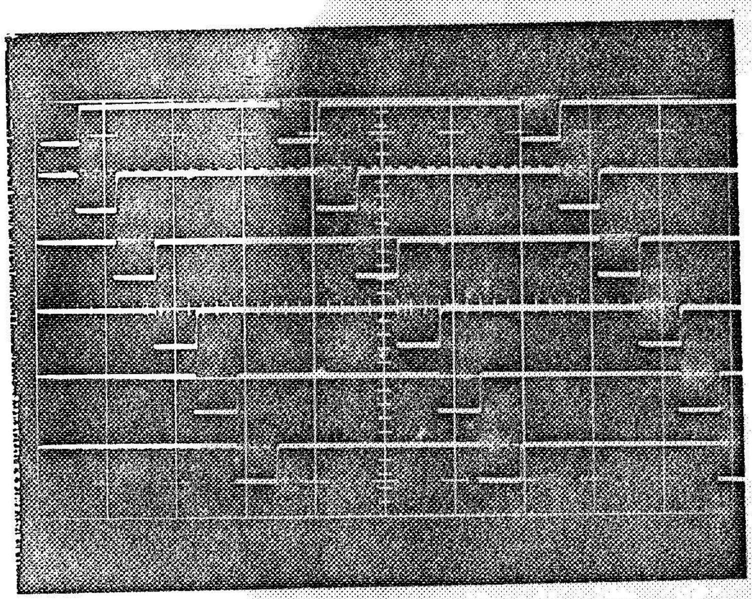
B

C

D

E

F



200 microseconds per cm

A first stage
B second stage
C third stage
D fourth stage
E fifth stage
F sixth stage

all 10 v/cm

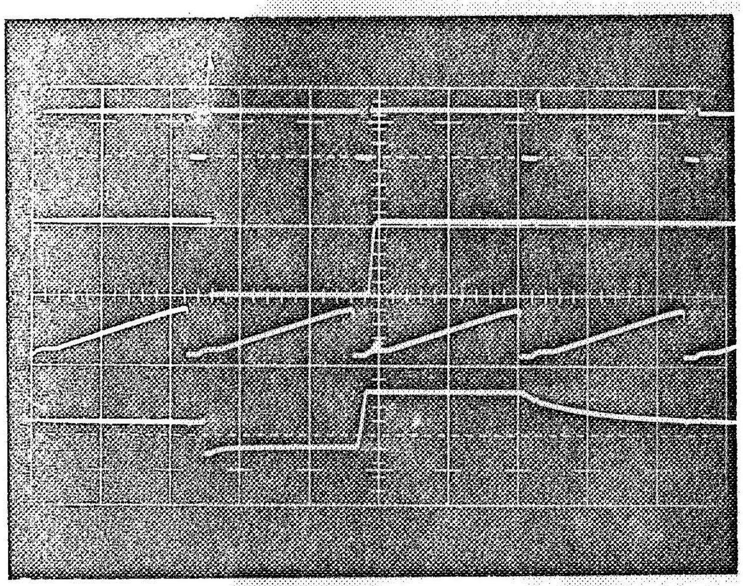
Figure 6.12 Transmitter ring osc stage outputs

A

B

C

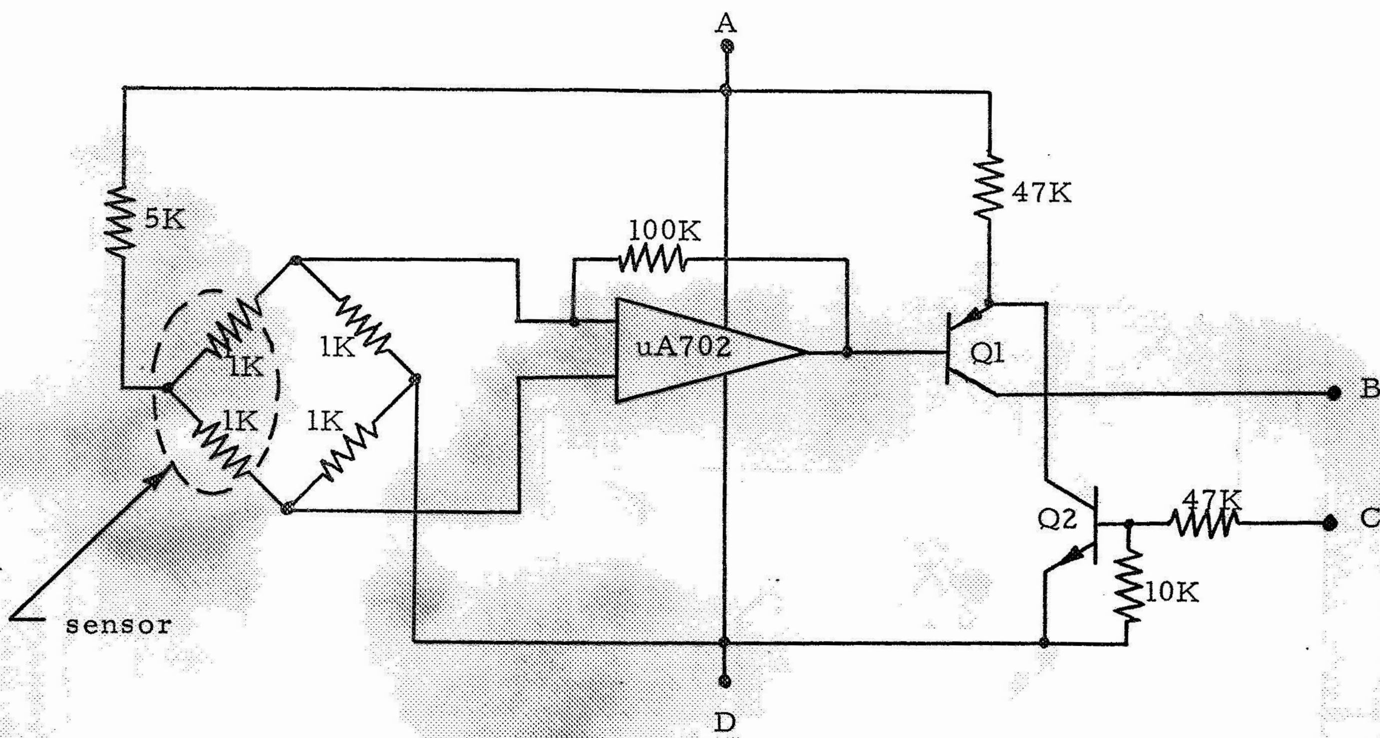
D



50 microseconds per cm

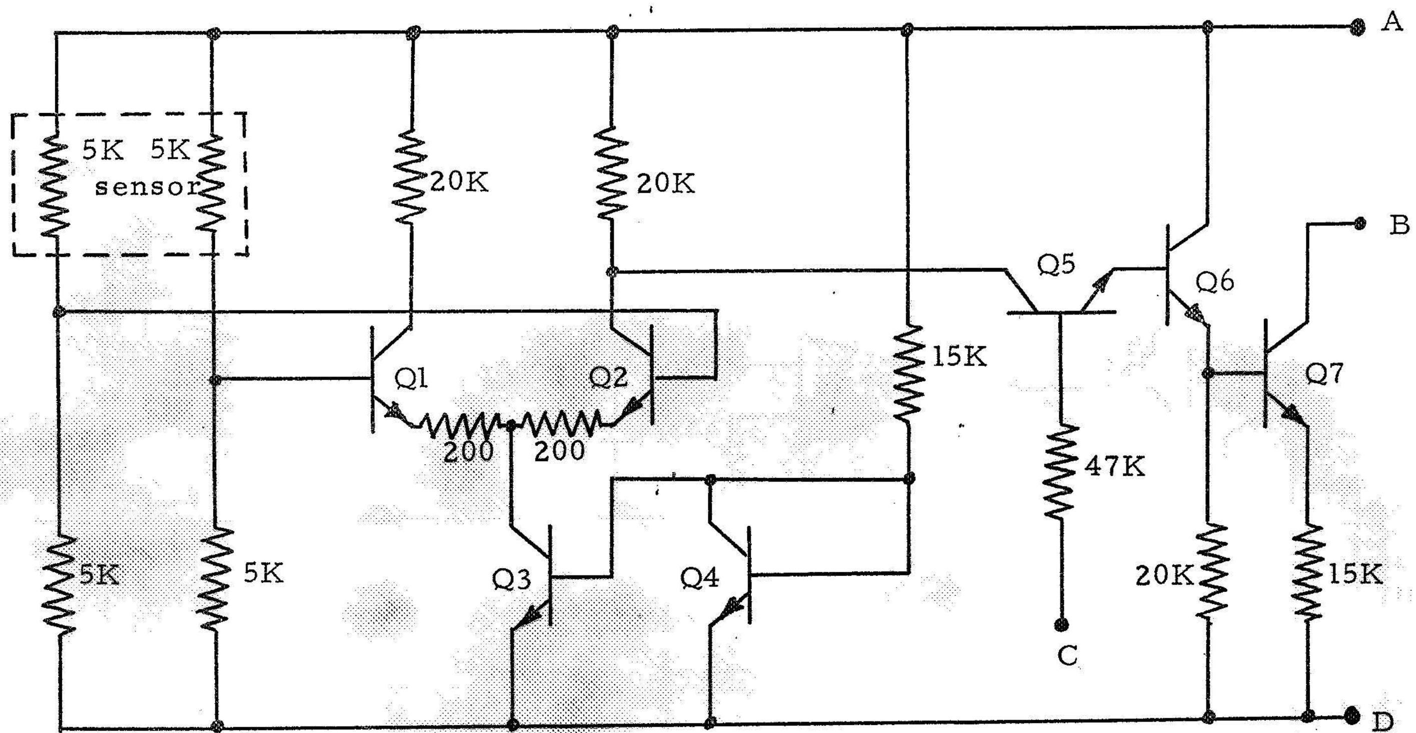
A trigger line 2v/cm
B first ring osc stage output 5v/cm
C ring osc driver SCS anode voltage 5v/cm
D second ring osc stage SCS cathode gate voltage 2v/cm

Figure 6.13 Detailed transmitter waveforms



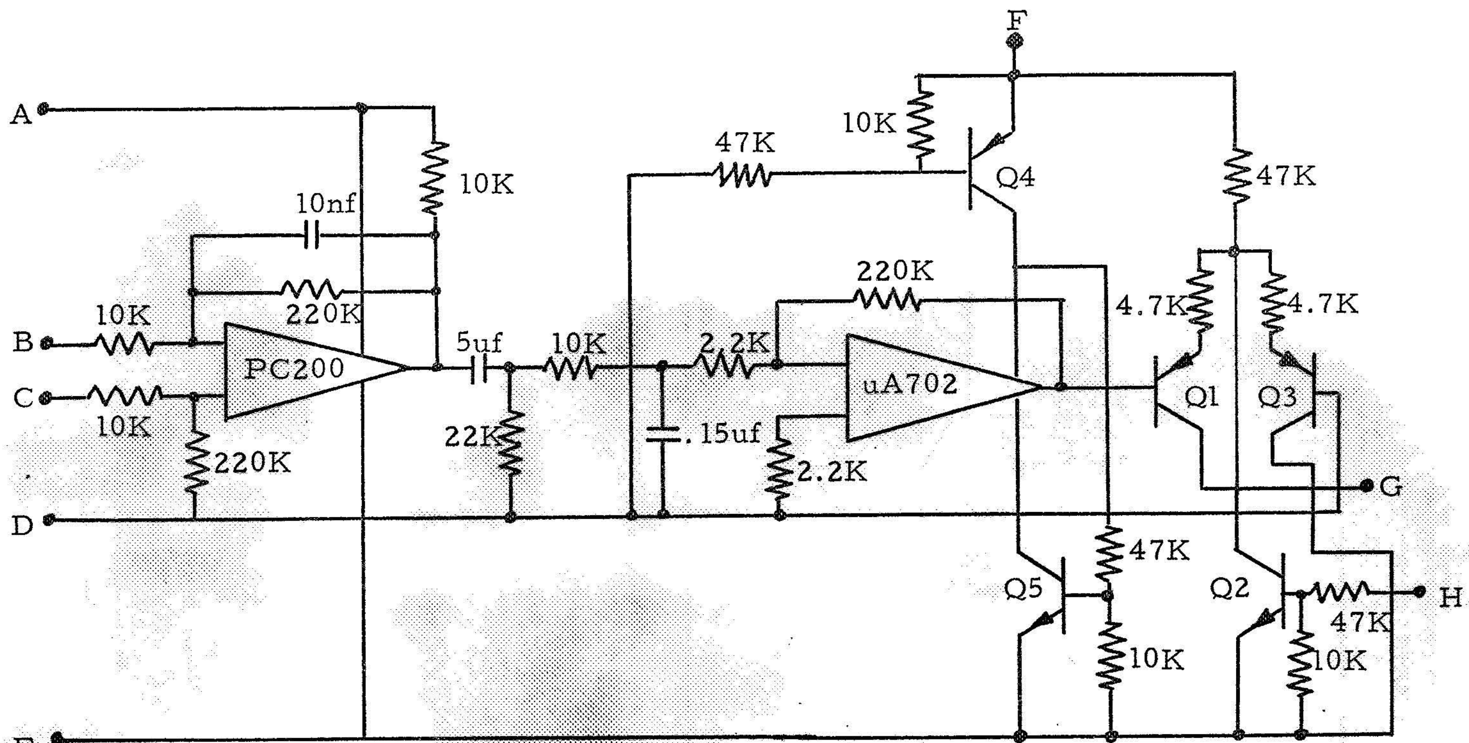
A) ring osc stage output | C) next ring osc stage output | Q1 2N3638
B) PAM signal output | D) negative supply | Q2 2N2923

Figure 6.14 Resistance-bridge channel input amplifier



A) +5 volt supply
B) PAM output to modulator
C) next ring osc stage output
D) ring osc stage output
all transistors 2N3663

Figure 6.15 Revised resistive bridge input amplifier



A) positive supply
B) +C) input signal
D) common

E) negative supply
F) ring osc stage output
G) PAM signal output

H) next ring osc stage output
PNP 2N3638
NPN 2N2923

Figure 6.16 Electrical-signal channel

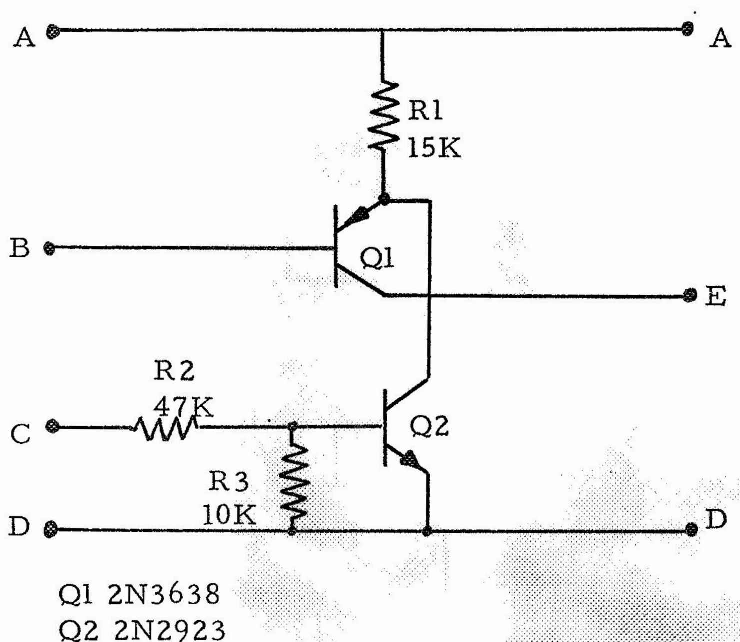
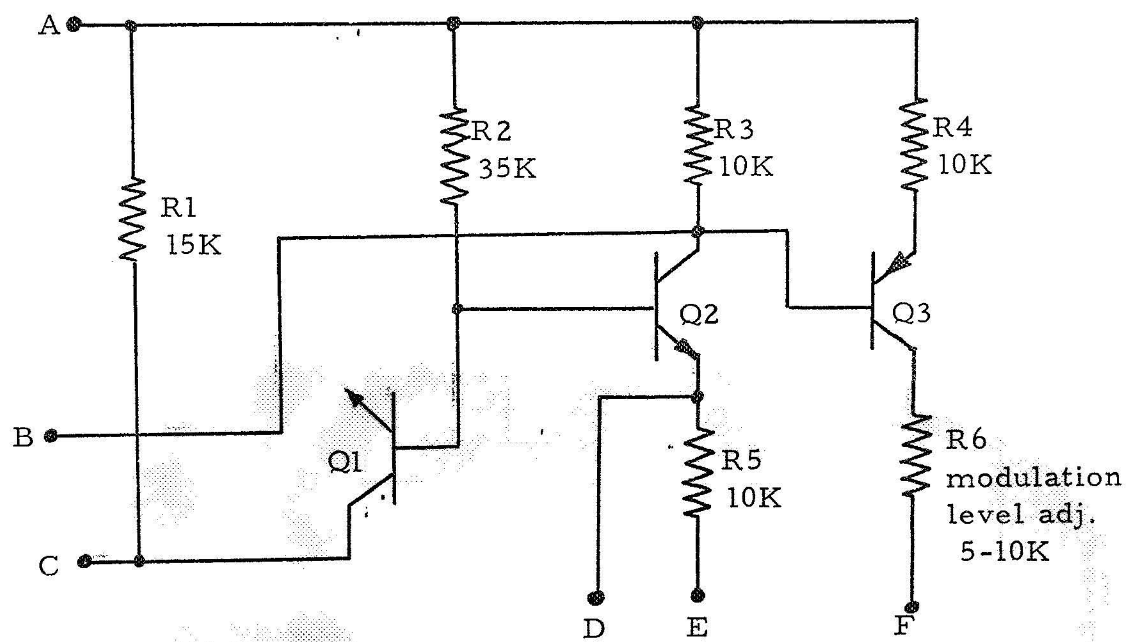


Figure 6.17 Frame-sync-signal generator

A) Ring-oscillator output
B) O volts
C) Ring-oscillator output, following stage
D) - 2.5 volts
E) Output (PAM line)



A) +5 volts
B) PAM line input
C) second ring osc stage output
D) sync level adj.

E) first ring osc stage output
F) common
Q1 + Q2 2N3663
Q3 2N3638

Figure 6.18 Modulator and sync level generator circuit

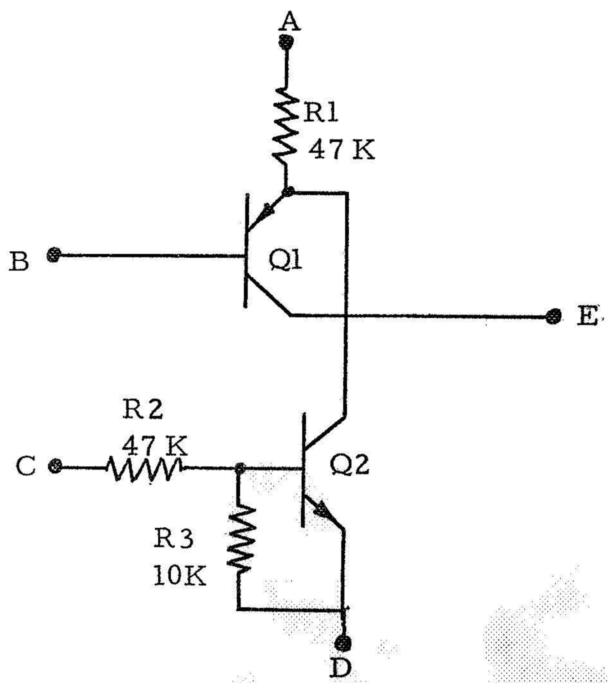


Figure 6. 19 Multiplexing gate

A) Control signal
B) Input
C) Control signal, following channel
D) Negative power-supply
E) Output

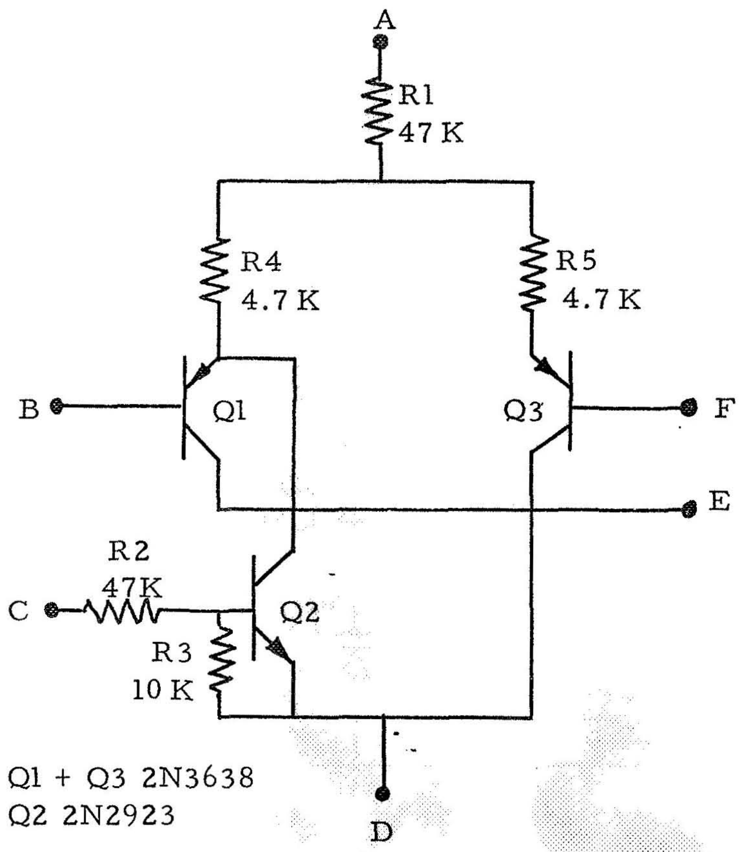


Figure 6. 20 Higher-gain multiplexing gate

A) Control signal
B) Input
C) Control signal, following stage
D) Negative power-supply
E) Output
F) Zero volts

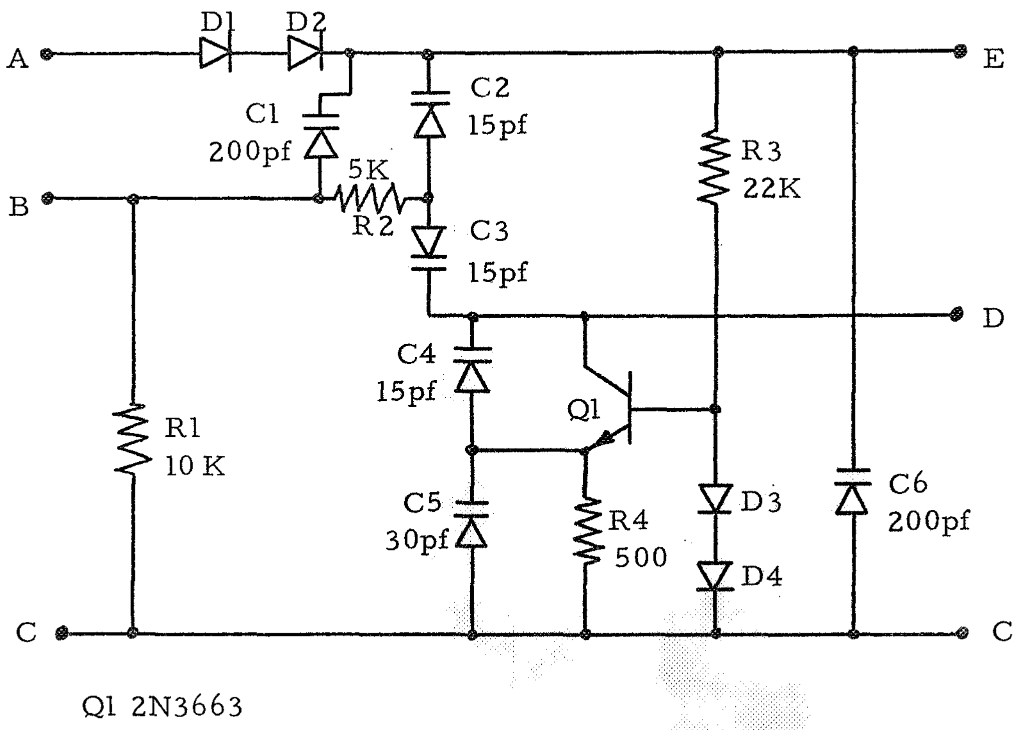


Figure 6. 21 RF unit

A) + 4 volts (either A or E used as positive supply input)
B) Modulating input
C) - 2. 5 volts
D) RF coil (between D + E)
E) RF coil, also decoupled + supply

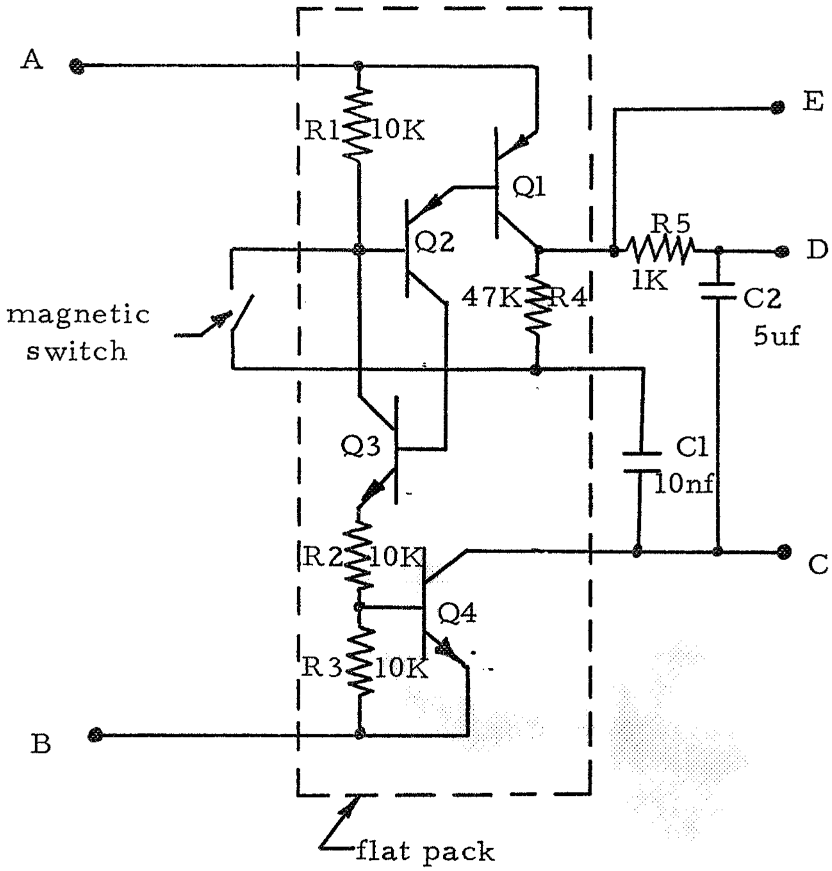


Figure 6.22 Remote-switch circuitry

A) + 4 volts, input
B) - 2.5 volts, input
C) -2.5 volts, output
D) Decoupled output for RF stage
E) + 4 volts, output

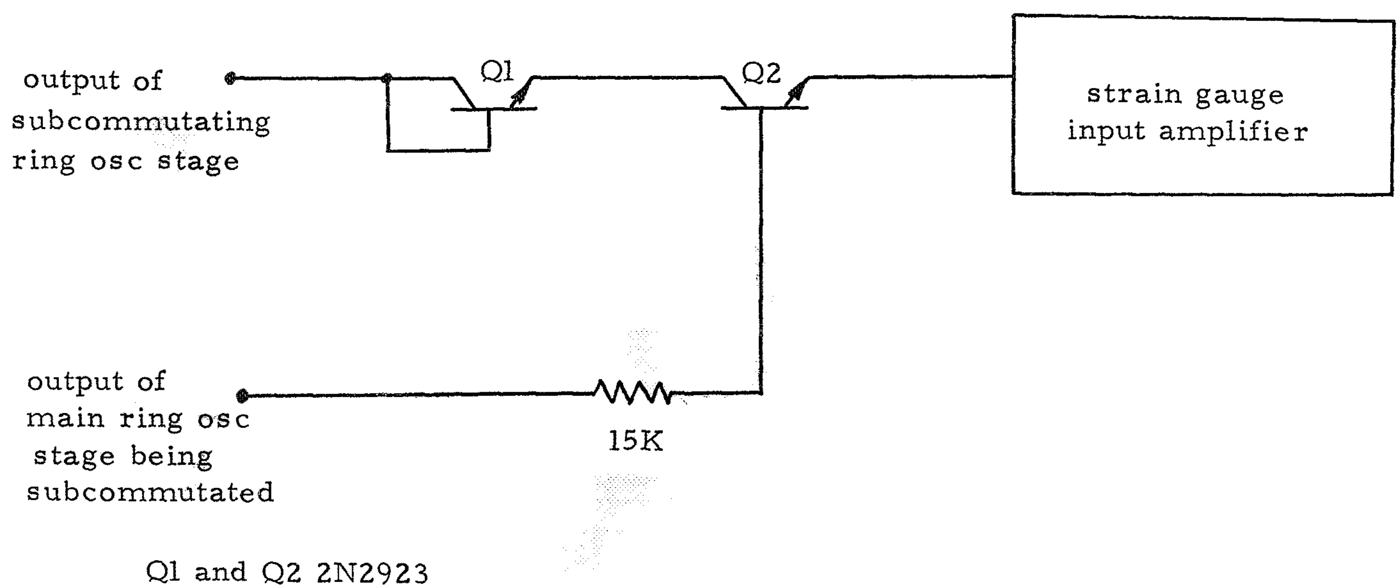


Figure 6.23 Modification of original strain gauge amplifier for subcommutation

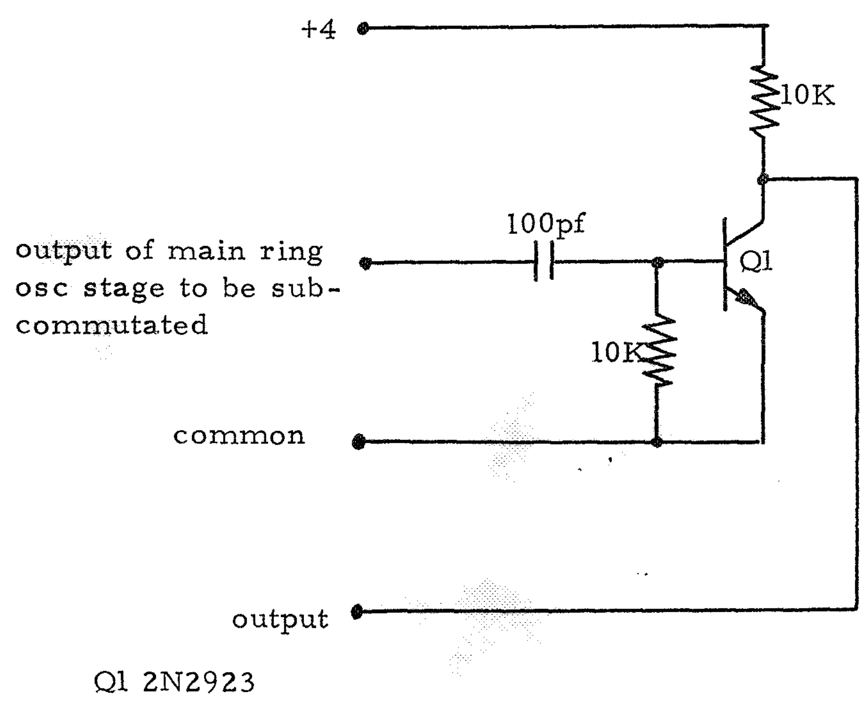


Figure 6. 24 Subcommutating ring oscillator driver

SECTION VII

DEMODULATOR DESIGN

The demodulator system performs a function which is the inverse to that of the multiplexing circuitry in the transmitter. Additional features are incorporated in the demodulator to give increased performance. The overall design hinges on making the transmitter as small as possible. The demodulator aids this by employing more extensive and complex circuitry to compensate for features which can be minimized or eliminated from the transmitter. The rule is to make the transmitter simple and utilize whatever circuitry in the demodulator necessary to give high reliability and performance.

The demultiplexing unit must accurately separate the received, composite-PAM, signal into the correct channels. This places the following requirements on the demodulating system:

a) Automatic gain control, AGC, for amplitude calibration

b) DC-level restoration or clamping for low-frequency response and elimination of low-frequency crosstalk

c) Synchronized desampling of the PAM waveform to assure the proper channel pulses to each output

d) Desampling after all switching transients have decayed to eliminate crosstalk between adjacent channels.

e) Output filtering to remove sampling frequency components from the outputs.

The demodulating unit is given in block-diagrams form in Figure 7.1. The operation of this unit falls into two main divisions. These are the steady-state operation and the turn-on and lock-in operation.

The demodulator uses a ring oscillator almost identical to that of the transmitter. Series diodes are used in the trigger inputs of the stages and diodes instead of transistors in the "analog and" circuitry to provide reverse breakdown voltages of 12 volts, the demodulator power supply.

The delays and windows for clamping and sampling use standard one-shot multivibrators. Special designs were developed for

a) sample-and-hold circuitry

b) AGC amplifier circuitry

c) clamp-level-sense circuitry

d) low-pass filters

that would perform the desired functions for the demodulator.

7.1 Steady-state operation

7.1.1 Ring oscillator

A ring oscillator, similar to the one in the transmitter, controls the demodulator operation. This ring oscillator is slaved to the received composite-PAM waveform. A sync-pickoff stage recognizes the frame-sync pulse and the ring oscillator is reset from this stage's output. The ring counter is then switched to the next stage by the trailing edge of the recognized sync-pulse. The switching period of the ring-oscillator driver is then locked to the PAM signal by an automatic-frequency-control (AFC) circuit.

7.1.2 Sync recognition and clamp level sense

This circuitry must recognize the sync pulse in the composite PAM and not be mis-triggered by noise pulses. The circuitry in Figure 7.2 performs this task. The first differential amplifier or comparator looks at the PAM and a reference level, and gives a positive output when the PAM exceeds the reference. The low-pass filter between the two comparators integrates the output of the first and the second compares the filter's output with a reference. The diode in the filter discharges the capacitor rapidly at the end of the sync or noise pulse. The time-constant of the filter will not allow noise pulses to integrate to a value large enough to trigger the second comparator. Thus, noise rejection is obtained. An inverted high-level output is also provided.

7.1.3 Sample-window generator

This generator provides the small desampling aperture discussed in Section IV and shown in Figure 4.1. The generator is triggered by the ring-oscillator driver so that the aperture is locked to the ring oscillator's switching times.

7.1.4 Sample-and-hold circuitry

Desampling a PAM waveform requires a sample-and-hold action. Pulses in the composite PAM for a given channel must be gated to that channel's output. The pulses should be stretched into 100 percent duty-cycle pulses to provide more energy to the input of the low-pass filter which is at the channel's output.

The sample-and-hold circuitry is shown in block-diagram form in Figure 7.3 The actual circuitry is given in Figure 7.4. The comparator looks at the output and the PAM waveform. The gated amplifier is controlled by the "and" combination of the sample window and the ring-oscillator-stage output corresponding to that channel, and is driven by the comparator's output. This amplifier adjusts the charge on the capacitor so that the input and output are equal during the gating period. Between gating periods the capacitor sees a very high impedance load so that it maintains its charge during this interval. The gated amplifier has a current-source output with infinite output impedance when gated off. The amplifier following the storage capacitor has a high input impedance and isolates it from the load on the channel's output. These impedances provide the long time constant for the holding operation.

This circuitry provides narrow sampling apertures, adjustable from 100 μsec to about 3 μsec as a limit, and holds the sample value for several milliseconds. This circuit is useful in several places; the most common is the demultiplexing gates. It is also useful as a clamp-drive circuit, adjusting the charge on the capacitor at the output of the AGC amplifier to provide clamping action on the PAM signal. It also is used to provide the AGC drive.

Low-pass filtering is used to remove the sampling-frequency components from the output and to provide the bandwidth limiting necessary for noise considerations.

7.1.5 Low-pass filters

Sharp-cutoff filters are available using inductors and capacitors. These are not suited to this project as the very fast risetimes of the sample-and-hold circuits make the filters ring, giving a high noise level in the outputs. For these reasons it was necessary to design a filter with a steep slope and no ringing.

In channels with low-bandwidth relative to the sampling rate, a three section resistor-capacitor filter was used as shown in Figure 7.5. This can be included on a printed-circuit card and performs well, for example, in 10 Hz channels with 500 Hz or higher sampling frequencies.

Channels using the minimum four-to-one sampling ratio require sharper filters. Figure 7.6 gives an active filter for this application.

7.1.6 Clamp and clamp-window

The top of the frame-sync pulse is locked or clamped to a fixed level by the clamp circuitry. The preceding amplifier has a capacitor-coupled output. The charge on this capacitor is adjusted by the clamping circuit to set the proper DC-level at the amplifier's output.

The clamp-window generator provides a narrow aperture for the clamping action after the switching transient has decayed and before the end of the sync pulse. This circuitry is triggered by the second sync-recognition comparator. This aperture is also the signal used to reset the ring oscillator to the stage corresponding to the frame-sync pulse.

7.1.7 Automatic gain control

The automatic-gain-control (AGC) circuitry consists of a variable-gain amplifier, a level comparator and a low-pass filter. The AGC pulse is compared with a fixed level by the level comparator. The comparator provides a signal to adjust the gain of the amplifier. The low-pass filter is included to keep the AGC loop stable. The AGC drive is gated by the combination of the sample window and the ring-oscillator-stage output corresponding to the AGC pulse.

7.1.8 AGC amplifier

The AGC amplifier is a variable-gain wide bandwidth amplifier. A photocell-lamp combination is used in a variable attenuator circuit. This is the only linear wide-signal-swing variable impedance element available which is non-mechanical. Operating the lamp at half of rated voltage or less will give it an indefinite lifetime. This combination has a response-time of about one second and this must be taken into account in constructing the AGC feedback loop to maintain stability. Figure 7.7 gives the circuit diagram for the AGC amplifier.

7.2 Turn-on and lock-in operation

Acquisition of locked operation upon system-turn-on and reacquisition, after disturbances due to noise or after other loss of lock, requires special characteristics in the circuitry. The composite PAM at the input to the sync-recognition circuitry must have the proper levels and amplitude to initiate clamping-action.

The AGC amplifier is designed so that its gain becomes a maximum when the frame-sync circuitry is not operating. This gives a maximum signal-swing to switch the input comparator of the sync recognition circuits. This starts the clamping action. A resistor to the positive supply from the capacitor-coupled output of the ACG amplifier causes the DC level at that point to drift in the positive direction. This, coupled with maximum AGC amplifier gain, assures that the PAM will swing past the sync-recognition level.

The AGC drive is inhibited until clamping action begins. At this point the AGC circuitry begins to function, reducing the gain to the proper level and the demodulator becomes locked in and ready for operation.

7.3 Improved Demodulation Circuitry

A block diagram of the multiplex receiver modification is shown in Figure 7.8. This circuitry uses the level clamp and AGC functions of the original circuitry. The clock frequency is adjustable and in operation it is adjusted to the clock frequency in the transmitter. The clock synchronization allows for minor frequency variations and assures a proper phase relationship between the two clocks. Thus, the receiver clock puts out a pulse each time a different channel appears on the incoming signal.

A General Instrument MOS Integrated Circuit, MEM 3012 SP 12-BIT serial in-parallel out shift register, is used to select the proper sample and hold circuit to sample the incoming channel. At the beginning of a cycle of channels coming in (as detected by the sync level detector) a data pulse is fed to the MEM 3012 SP. As each clock pulse comes along, the data pulse is shifted from one output to the next. The pulse appears first in the first output, then when a clock pulse arrives in the second output and so forth.

Each output of the shift register is fed to an "and" gate. A second input of the "and" gate is connected to a sample window pulse generator. This generator produces a pulse of controlled width in each period between clock pulses. The controlled width pulse is delayed a constant amount from the beginning of the period. The sample window pulse is delayed to allow any switching transients to die down before sampling occurs. The sample window pulse widths may be as short as 1 microsecond.

The output of each "and" gate is fed to the gate of the corresponding sample switch. General Instrument MEM 2009 MOS I. C. multiplex switches are used for the sample switches. The MEM 2009 consists of 6 MOS FET switches. The source connections of all FET's are tied together. The signal line is brought in on this connection. As each gate is turned on, the FET becomes a low resistance and connects the signal line (source) to the hold capacitor (drain). The signal in the line at that time is stored in the hold capacitor. Another MEM 2009 is used to read out the voltage stored on the capacitors. The FET's are used as source-follower (unity gain) amplifiers. The source's of the FET's which are internally tied together are held at - 6 volts, the drains are individually fed through load resistors to + 12 volts. The gates are connected to the 'hold' capacitors. (Note: The drain and source are reversed from usual source follower connections, but because of the symmetry of an FET this does not matter.) The high input impedance of the MOS FET insures that little charge will leak off the 'hold' capacitor.

Each MEM 2009 flat pack contains six MOS FET's. At least two flat packs must be used together. A total of four flat packs will provide for up to twelve channels of information. The MEM 3012 SP shift register also provides for twelve bits of information. Thus, any number of channels up to twelve may be demodulated.

7.3.1 Clock Circuit Description

The diagram for the clock circuit is given in Figure 7.9. The clock consists of a simple unijunction transistor relaxation oscillator with a pulse shaper and a synchronization circuit. The pulse shaper is required to obtain adequate pulses for operating the shift register. The synchronization circuit differentiates the sync level pulse, providing a sharp spike. This spike turns on a pnp transistor which shorts the charging resistor and fires the UJT.

The frequency of the clock may be varied by adjusting the 50 K potentiometer. The clock frequency may be varied over an even wider range by changing the timing capacitor. The period _must_ be adjusted such that the synchronization pulse does not cause two clock pulses to appear very close together.

7.3.2 Delay and Sample Window Generation Circuitry

The delay is generated by an adjustable monostable multivibrator. The delay monostable triggers another adjustable monostable which provides a delayed sample window pulse. The delay is generally set to be one-half of a clock period. The sample window pulse may be as short as 1 microsecond and still allow the sample and hold circuit adequate acquisition time.

7.3.3 Sync Level Detector Circuitry

The sync level detector is the same as used on the original receiver.

7.3.4 Data Monostable Multivibrator

An ordinary adjustable monostable multivibrator is used to provide a data pulse as an input to the shift register when triggered by the sync level detector. The monostable multivibrator must be adjusted to provide a pulse for longer than one clock period, but shorter than two periods.

7.3.5 Shift Register

The shift register is a General Instrument MEM 3012 SP MOS I. C.

7.3.6 "And" Gates Circuitry

The "and" gates are simple TTL gates. The first pnp transistor from the shift register output provides isolation for and inversion of that output. Speed up capacitors are used to provide fast rise times for microsecond sample window pulses. The diagram for the "and" gates is given in Figure 7.10.

7.3.7 Sample Hold Circuit

The sample and hold circuit as given in Figure 7.11 makes use of two MOS FET's contained in separate MEM 2009 MOS I.C.'s. The first FET acts as an analog switch connecting the signal line to the hold capacitor upon command from the "and" gates. The hold capacitor stores the signal until it is updated. The second FET acts as a unity gain amplifier. Its high input impedance (10^{10} ohms) allows the hold capacitor to maintain its voltage level until the next sample window for that channel. The amplifer FET is operated as source follower (with source and drain levels inverted).

Figure 7.12 gives voltage waveforms for the new demodulator circuitry which were taken during the reception and demodulation of a 4 channel PAM signal.

7.4 Subcommutated Channel Demodulator

The receiver demodulator circuitry for this subcommutating system is basically the same as that for the main system except that no AGC function is necessary. A sync level somewhat lower than the main system sync level is used to determine the starting point for the subcommutating receiver ring oscillator. A block diagram of the subcommutating demodulation system is shown in Figure 7.13.

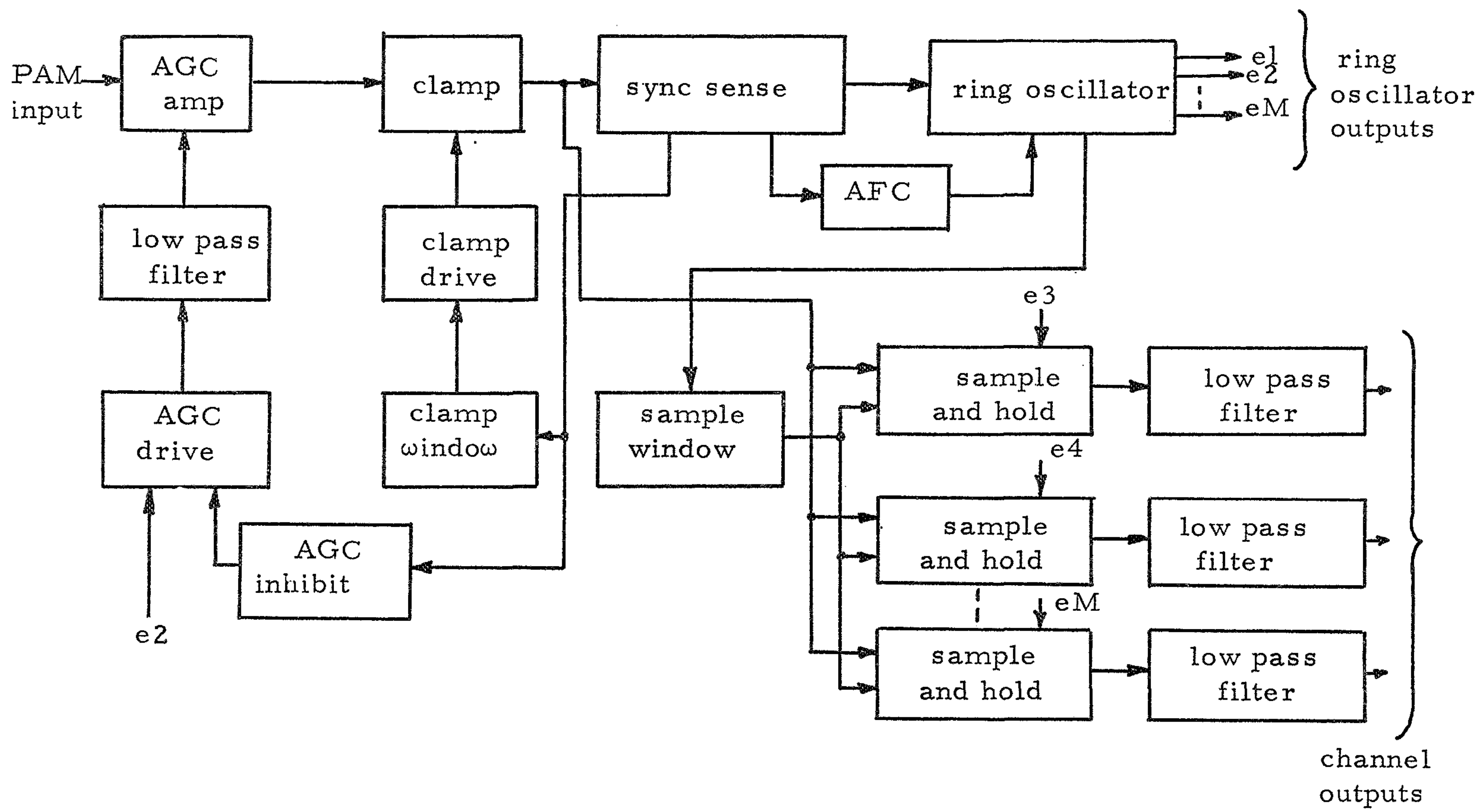
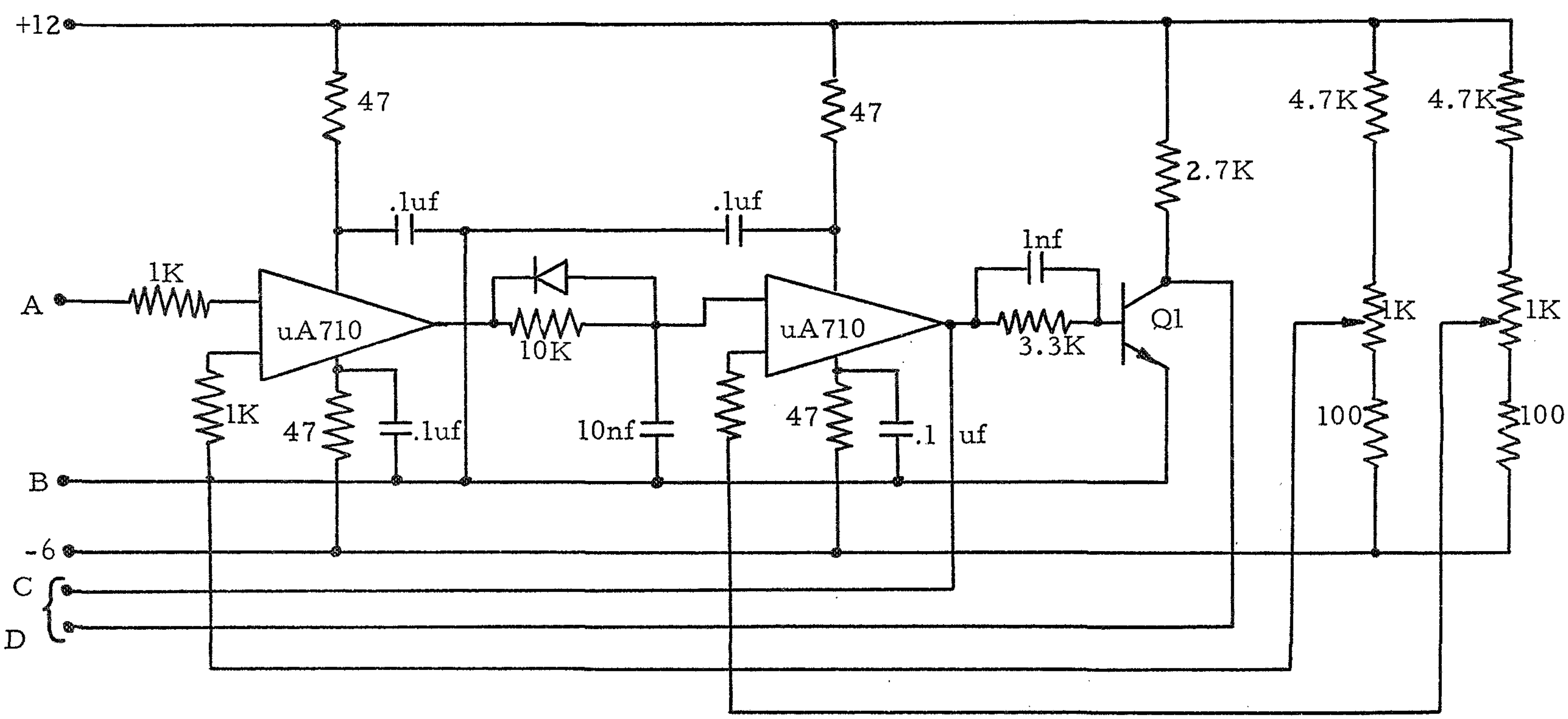


Figure 7.1 Demodulator block diagram



Q1 2N2923

Figure 7.2 Clamp level sense
A. PAM signal input B. common C+D. outputs

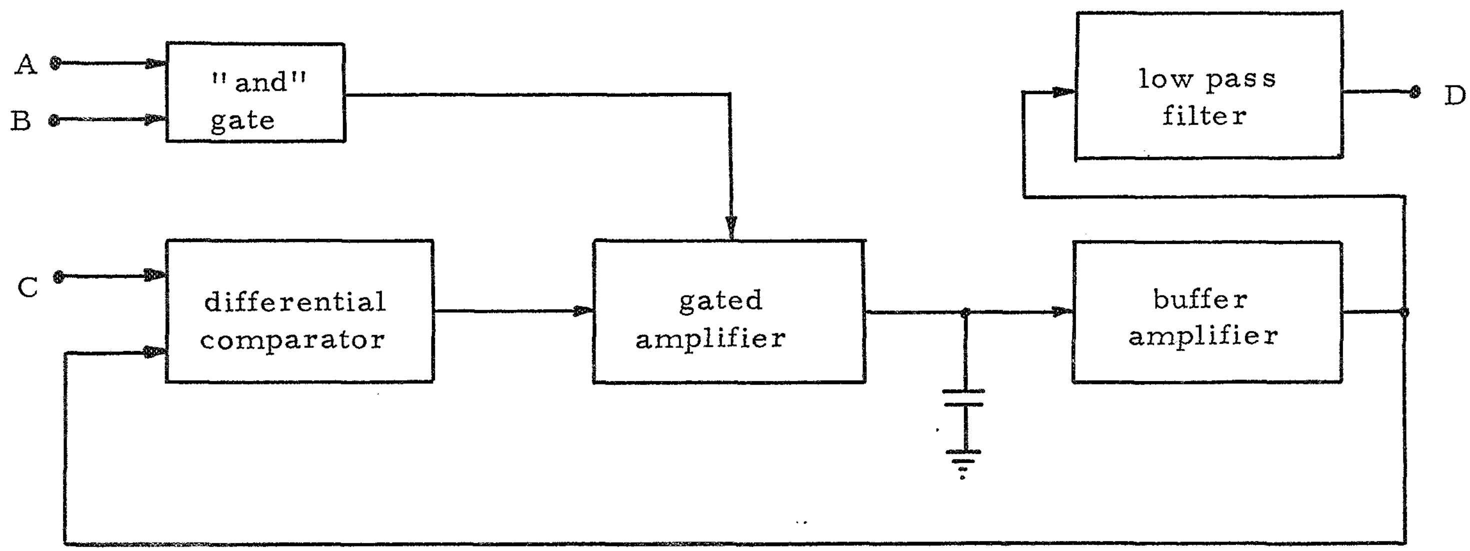
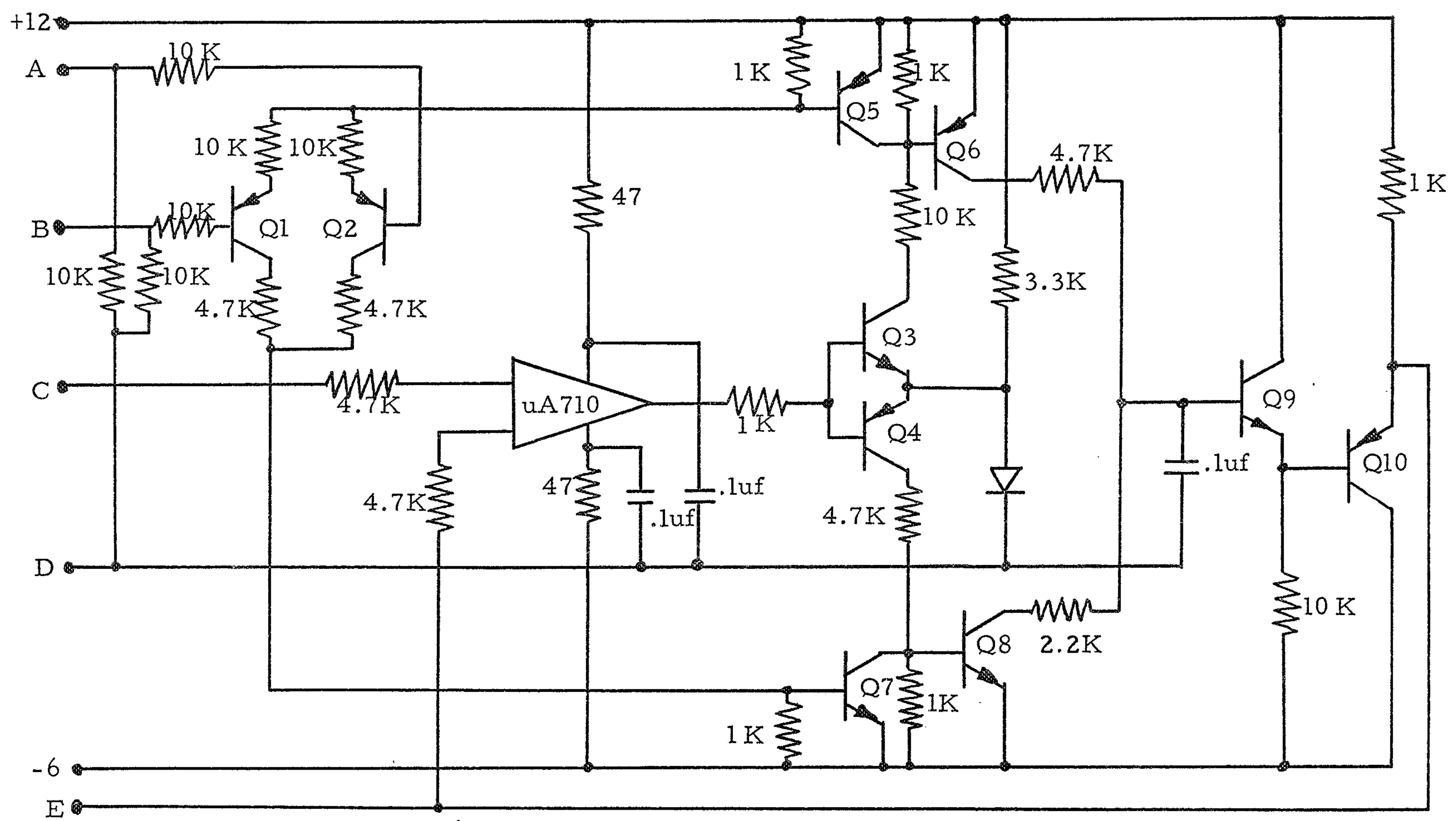


Figure 7.3 Block diagram of sample-and-hold circuitry

A and B) Gating inputs
C) Signal input
D) Output



PNP 2N3638
NPN 2N2923

Figure 7.4 Sample-and-hold circuit

A. ring osc stage B. sample windows C. PAM signal input D. common E. demodulated output

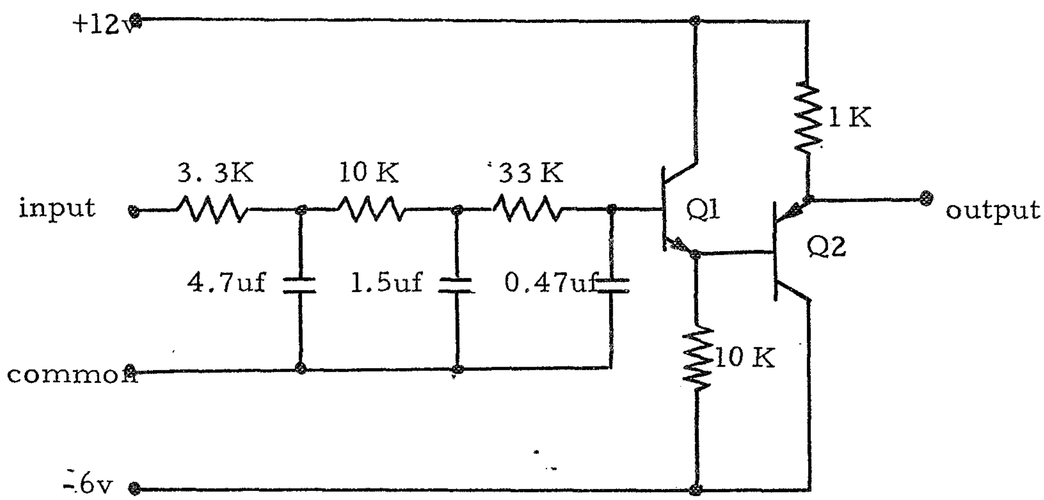


Figure 7.5 Low-pass filter

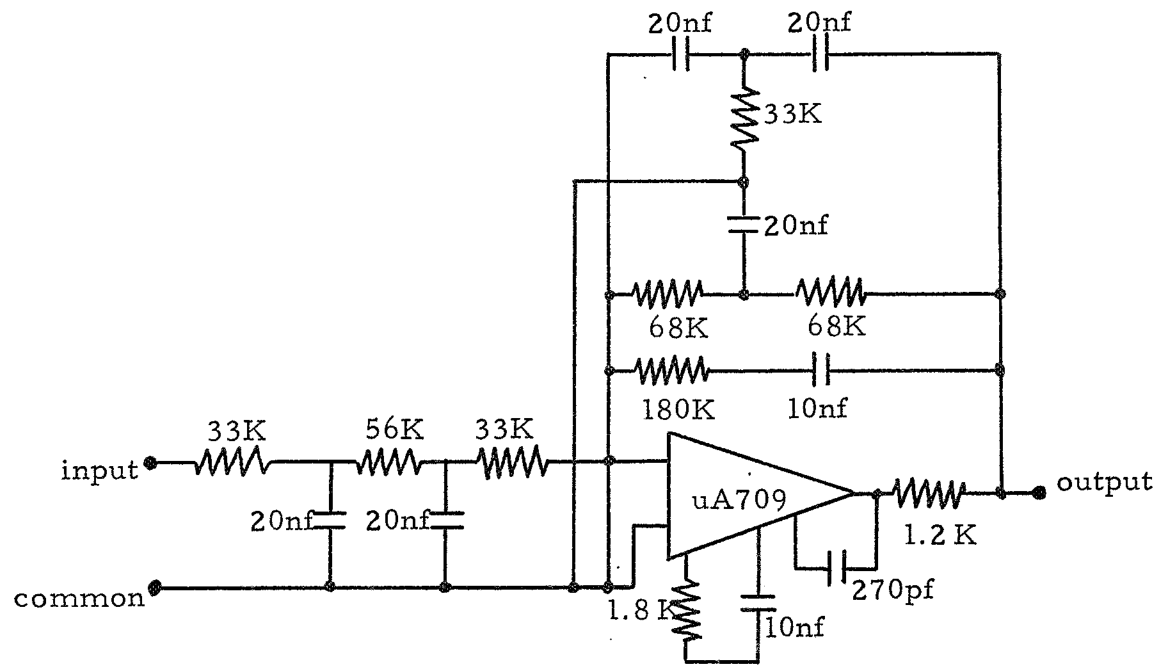
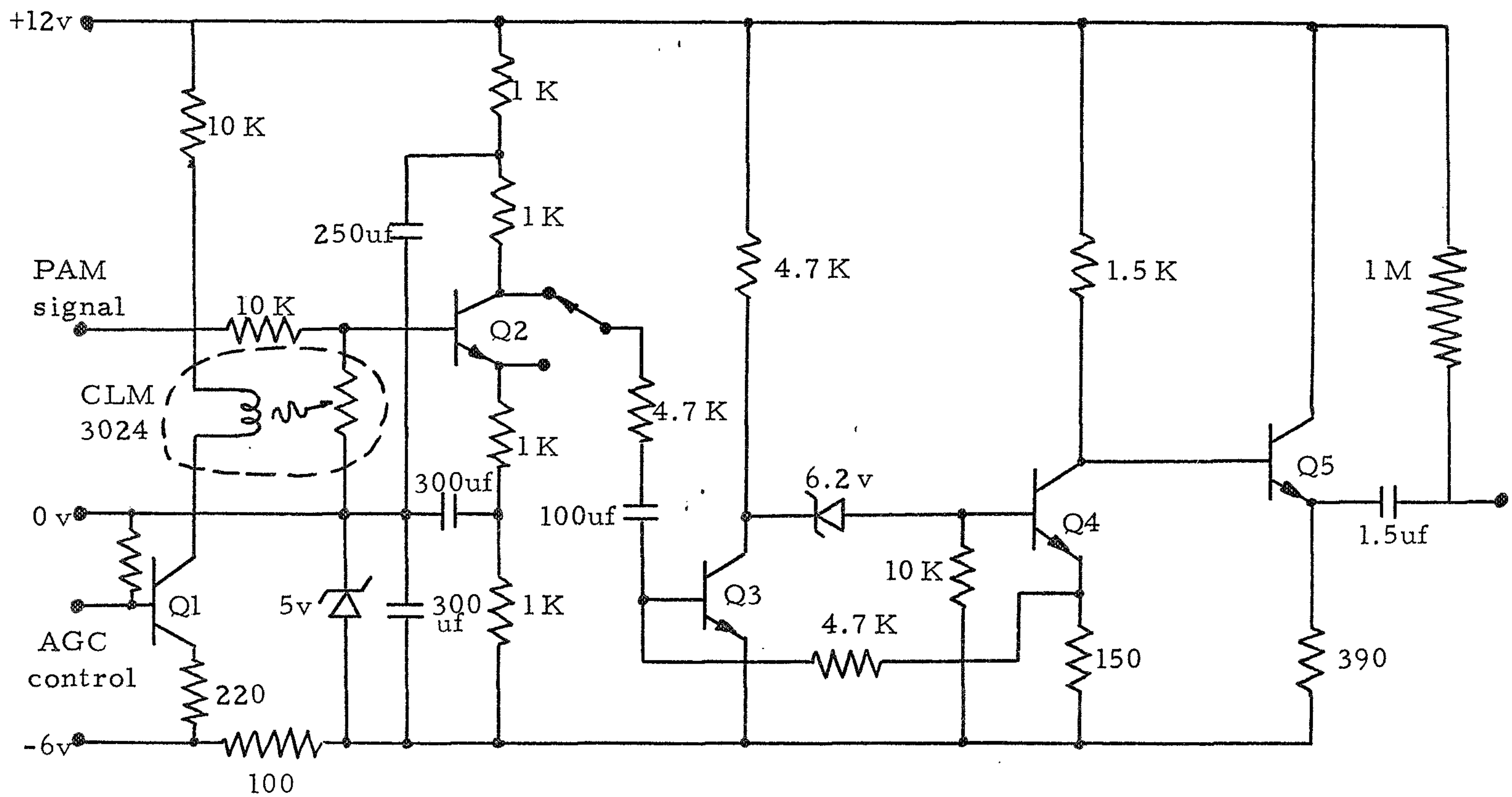


Figure 7.6 Active low-pass filter



all transistors 2N697   Q1 and Q5 require heat sink

Figure 7.7   AGC amplifier

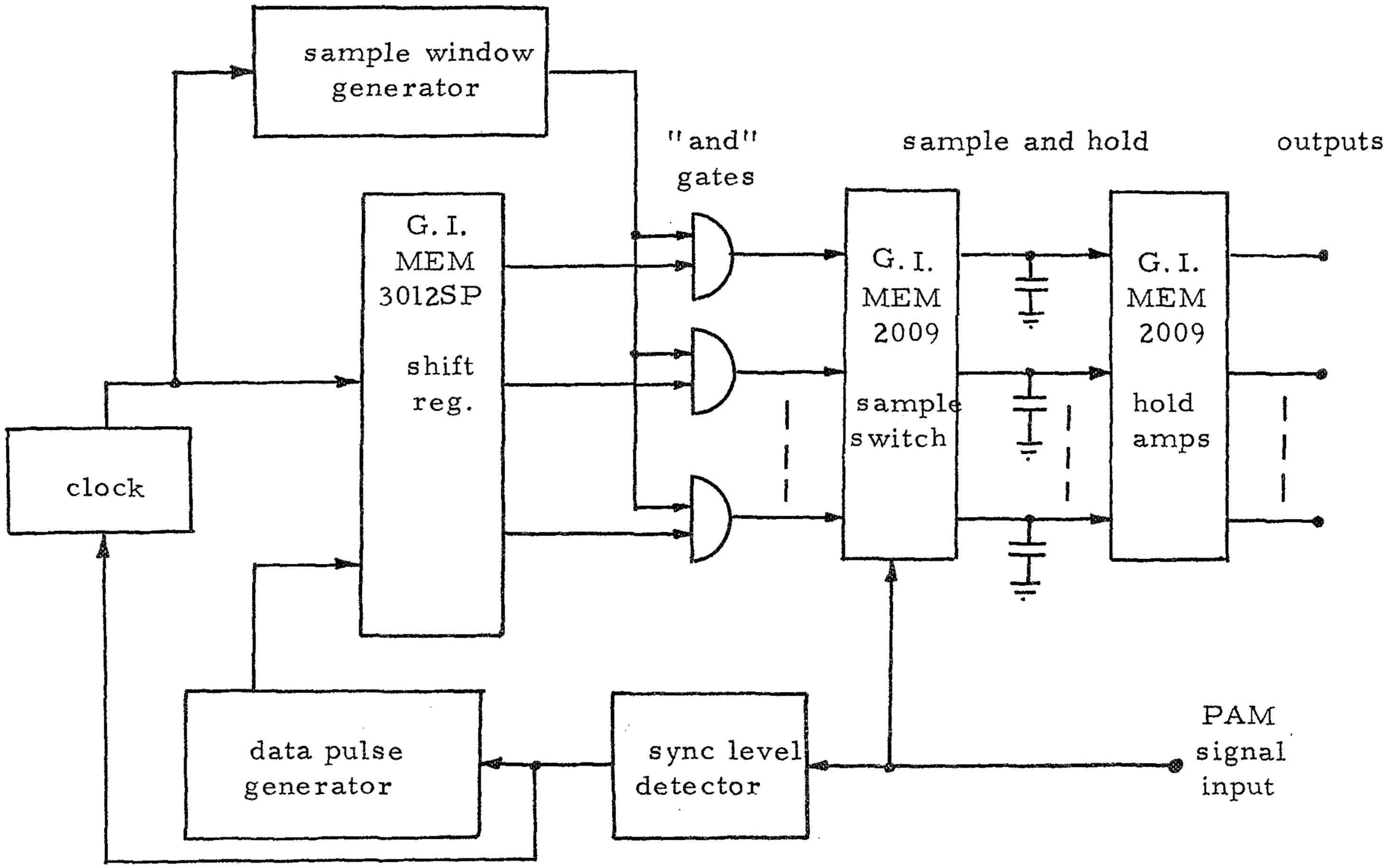


Figure 7. 8 Block diagram of improved demodulator

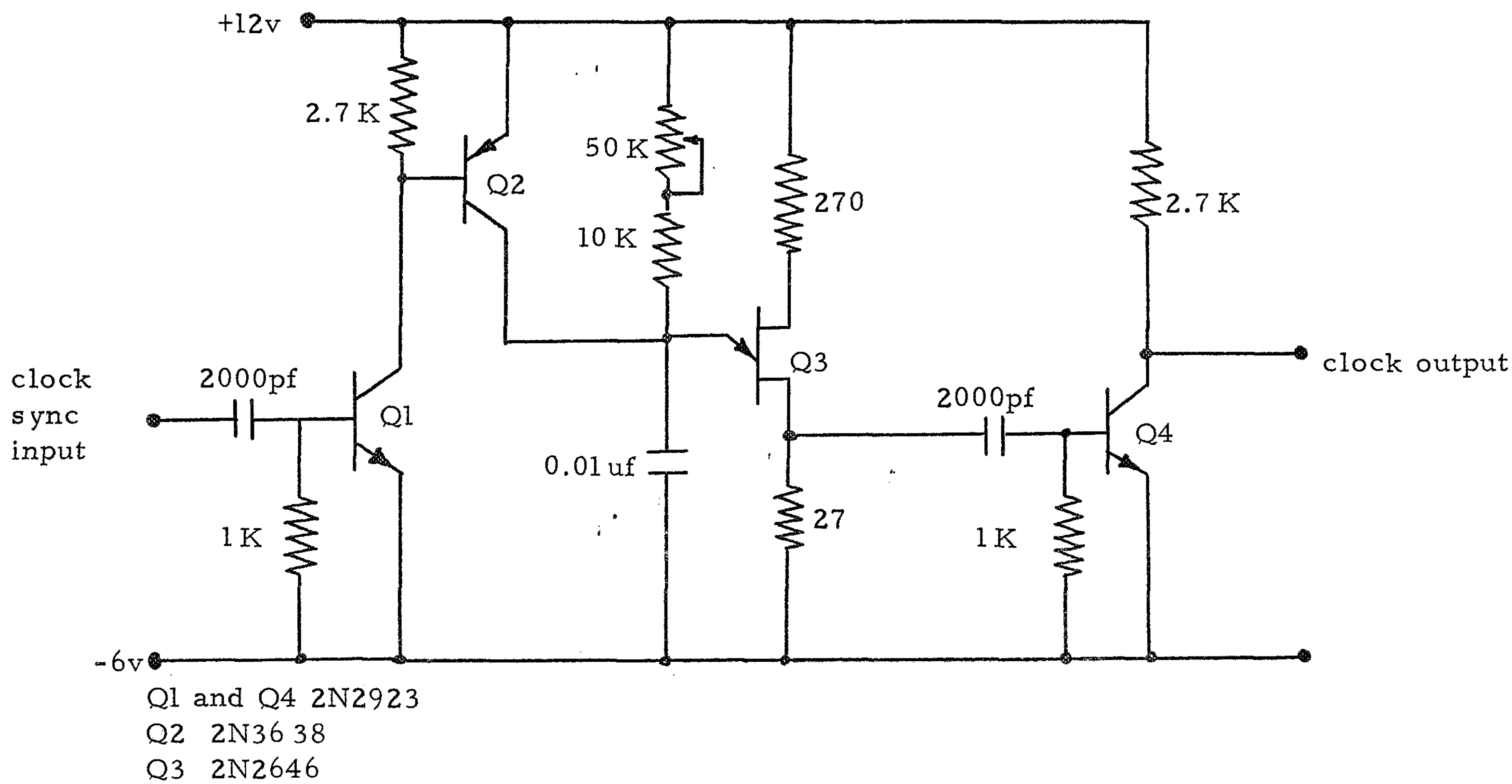
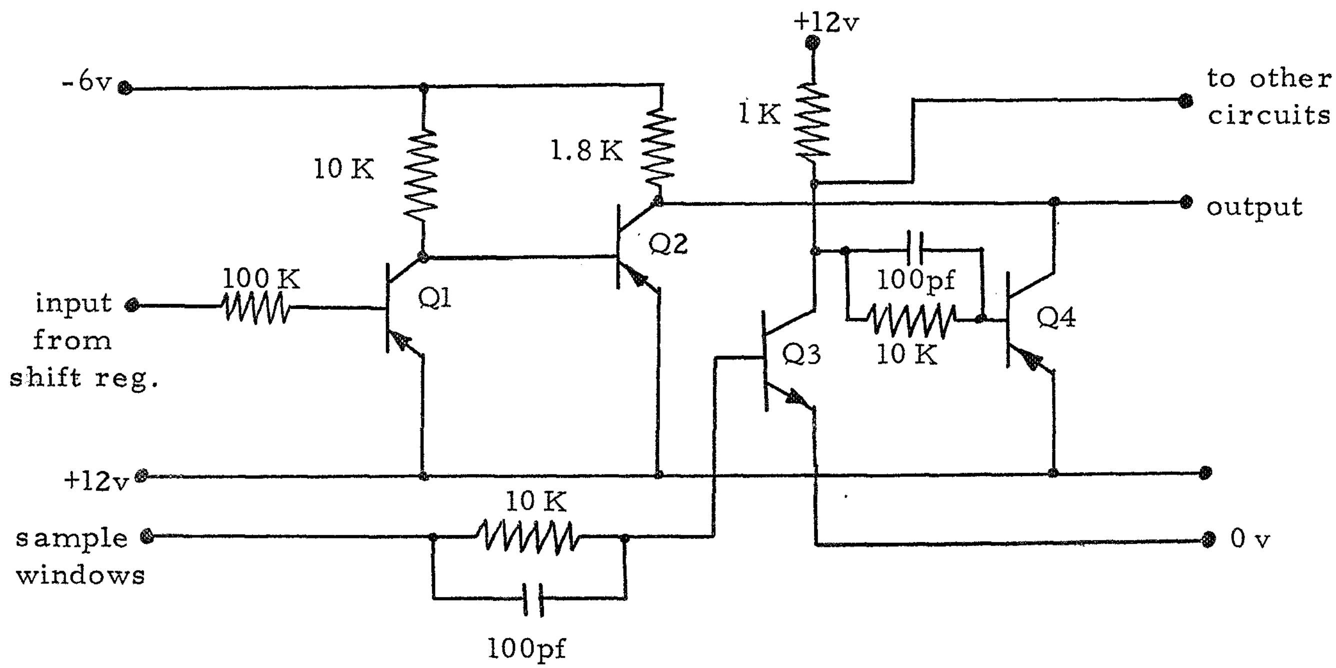


Figure 7.9 Revised clock circuit



PNP 2N3638
NPN 2N3663

Figure 7.10 "And" gates

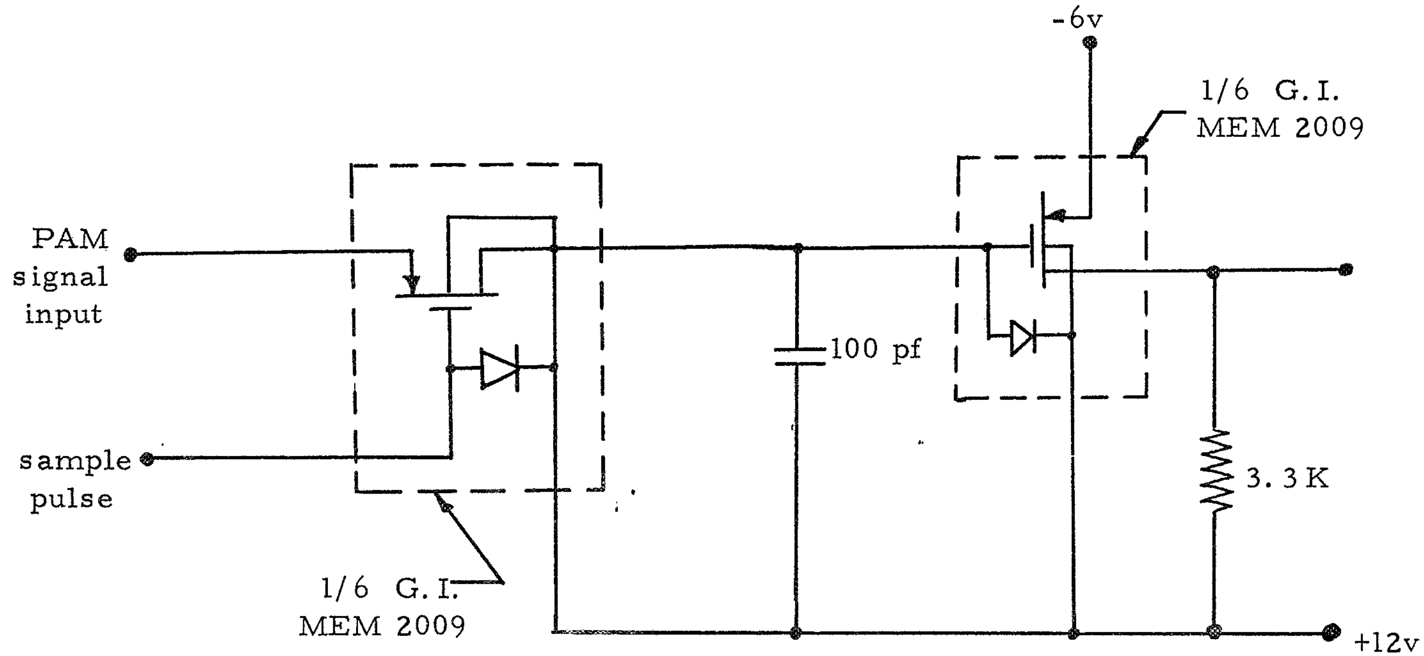
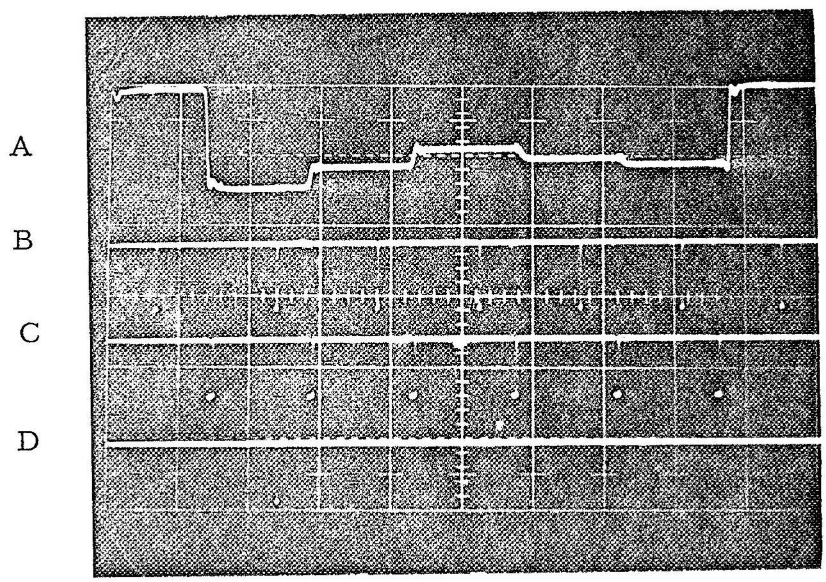


Figure 7.11 A single sample-hold circuit



100 microseconds per cm

Figure 7.12 Demodulator waveforms

A. Received PAM signal 2v/cm

B. Output of sample window generator 10v/cm

C. Demodulator clock output 20v/cm

D. Sample pulse for first channel 20v/cm

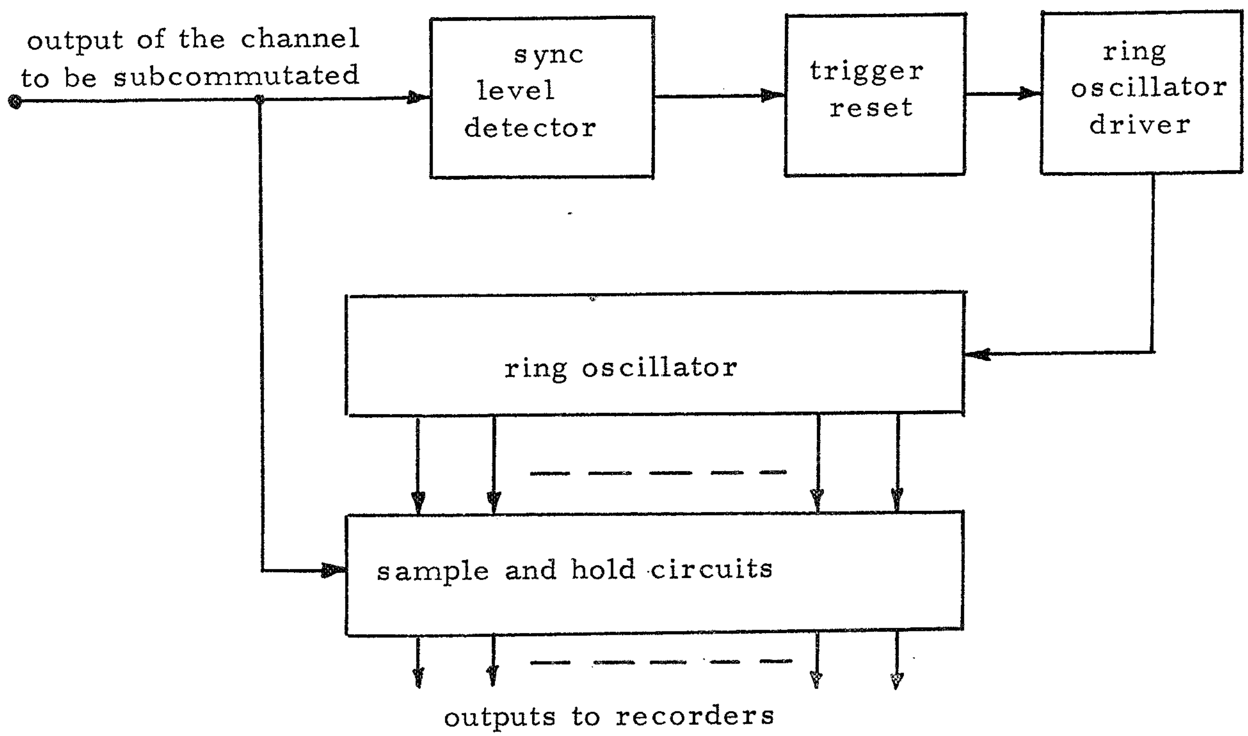


Figure 7.13 Block diagram of subcommutating demodulator

SECTION VIII

SYSTEM PERFORMANCE

A four channel discrete component transmitter was built and tested. Testing included implant in the abdominal cavity of a dog. A four channel hybrid integrated circuit transmitter was built, using circuitry identical to that of the discrete component transmitter. Difficulties encountered in the assembly and testing of this unit indicated some revision of the circuitry was necessary. Redesign of the transmitter ring oscillator has cleared up the problems which presented themselves in the previous design. A four channel hybrid integrated circuit transmitter using the new circuitry was built and tested. No problems were encountered in the assembly and testing of this transmitter.

An eight channel transmitter using the original circuitry was breadboarded and it successfully telemetered EKG, EEG and strain gauge information with no perceivable crosstalk. The telemetered records were identical to those obtained using a hard wire system.

8.1 Implanted transmitter

A four channel transmitter using micro-miniature discrete components was implanted in a dog. This implant was performed at the Pharmacology Department, University of Michigan. Four

1000 Ω-silicon strain-gauge transducers were used. These were sutured to the gastral antrum and duodenum of the dog to record contractite activity. The tests were performed with the dog and the receiving system's antenna in the same room, transmission distances being 20 feet or less. A continuous 24-hour record was obtained from the unrestrained dog, possibly the first record of its kind made anywhere.

The transmitter had a peak frequency-deviation of 200 kHz, RF bandwidth of 1 MHz and video bandwidth of 100 kHz. The sampling frequency was 1200 Hz per channel. Filters in the demodulator outputs limited the channels' bandwidths to 10 Hz for noise reduction. Full scale sensitivities were 500 gms with noise levels of 2 gms peak-to-peak. (less than 0.5 percent).

Activity patterns, telemetered by the implanted transmitter, are compared with patterns obtained from another dog direct-wired to recording equipment, in Figure 8.1. The two sets of recordings are very similar. Amplitude and activity patterns for similar transducer locations are nearly identical[9].

Figure 8.2 gives a picture of the ring oscillator, signal channels, multiplexing circuitry and transducers for the implant transmitter. The volume for this circuitry is 0.5 cubic inches. RF stage and remote switch add about 0.2 cubic inches. Battery pack for 100 hours operation and epoxy to match electronics-package shape to battery-pack shape bring the total volume to 2.5 cubic inches.

Integrated circuitry has reduced the volume for a similar four channel transmitter to under 0.5 cubic inches for electronics.

Power supply requirements for this transmitter were 35 milliwatts. Changing to 5000 Ω transducers from 1000 Ω units would lower the power requirements by about 5 milliwatts.

8.2 Eight channel breadboard system

Integrated circuit designs were tested in a breadboarded transmitter. Information channels were incorporated for resistance-bridge sensors and for direct-electrical signals. Initially, standard component circuitry was used. Integrated circuits were substituted into the transmitter as they were developed. All circuitry, except for the direct-electrical-signal channels, was reduced to integrated circuit form.

The tests were conducted in a laboratory. RF transmission distance was about 10 feet. Frequency deviation was 150 kHz peak. Video bandwidth was 100 kHz and RF bandwidth 1 MHz. The sampling rate was 600 Hz for the information channels, giving a useful information bandwidth of 150 Hz for each channel.

The electrical-signal channels had the following measured characteristics given in Table 8.1. The two channels use similar circuitry except for the multiplexing gates. The higher-gain channel uses the gate of Figure 6.18. The noise level of this channel is determined by the input stage. The lower-gain channel uses the gate given in Figure 6.17. Figure 8.3 gives calibration curves for these two channels.

EKG, picked up with surface electrodes, was telemetered using the lower-gain channel. Figure 8.4 gives a portion of this recording. Higher sampling rates would permit telemetering EMG with this channel.

EEG's, from electrodes implanted in the brain of a cat, were telemetered by two channels. A portion of these recordings is given in Figure 8.5.

Resistance-bridge channels were tested using 5000 Ω-silicon strain-gauge sensors. Figure 8.6 shows a sensor and the method of calibrating the sensors. Figure 8.7 gives a typical calibration curve for a strain-gauge channel through the system including demodulator. This sensitivity is higher than needed by the implant transmitter and will be reduced, resulting in lower power requirements in the channel. The noise levels were 0.5 percent of full scale peak to peak in a bandwidth limited to DC to 10 Hz in the desampling circuitry. Crosstalk was not measurable between any channels.

The power supply drain for the breadboard transmitter was 40 milliwatts. This transmitter included six channels for 5000 Ω strain-gauge transducers and two channels for electrical signals (EEG and EKG).

8.3 Hybrid Integrated Circuit Transmitter

A four channel strain gauge hybrid integrated circuit transmitter was built using the circuits which were tested in the eight channel breadboard transmitter.

It was found that the timing of the ring oscillator driver is directly connected to the overlap time between ring oscillator stages. This overlap time is detected in the "and" circuit which is fed back to the ring oscillator driver to reset the timing circuit. Although the overlap times only vary from one to six microseconds out of a frame width of almost 200 microseconds, this causes the frames to vary from one another by as much as 10%. This large variation is caused by incomplete reset of the timing circuit in the ring oscillator driver when the overlap time is less than five microseconds. Oscilloscope traces showing this variation are shown in Figure 8.8.

It was also found that the breadboard transmitter had introduced stray effects on the ring oscillator circuitry which were necessary for proper operation. This was found at assembly time when ring oscillator stages had to be carefully selected and matched to allow the ring oscillator to function.

By carefully selecting circuits, a transmitter was built and functioned properly. However, it was felt that the ring oscillator circuitry should be revised to minimize the possibility of stray effects causing a malfunction. A picture showing this completed transmitter is given in Figure 8.9.

The ring oscillator was redesigned using the same type of memory element, but combining the NPN-PNP combination into one unit, an SCS. Several additional advantages were accrued as a result of the ring oscillator revision. Two ring oscillator stages can now be wired in one flat pack instead of one. Only one polarity of power supply is required and the total power consumed by the ring oscillator is less than the previous design.

Sufficient hybrid integrated circuits were built and a four channel strain gauge transmitter was assembled using these circuits. No difficulty was encountered in either the assembly or testing of this unit and it is still functioning properly. Figure 8.10 shows a photograph of the completed transmitter and a printed circuit board identical to the one on which the transmitter was constructed. A scale is included for size comparison. The printed circuit board is approximately 3/4" x 5/8" and the height of the finished transmitter is 3/8". The power consumption of this transmitter is 15 milliwatts, 3 milliamps at 5 volts.

The linearity of the overall system using this transmitter is better than 2 per cent.

8.4 Subcommutating Breadboarded System

Transmitter and demodulator circuitry was breadboarded to allow subcommutation of one of the information channels. The system was set up to telemeter two high frequency channels and to subcommutate a third high frequency channel into four lower frequency information channels. Preliminary tests of this circuitry were successful, but the circuitry has not been put into hybrid integrated circuit form as yet. Results of the preliminary tests indicated no loss of crosstalk rejection or frequency response in the subcommutated channels with strain gauge input.

TABLE 8.1

MEASURED CHARACTERISTICS
OF ELECTRICAL-SIGNAL CHANNELS

Characteristic	Measured Value	
	Higher-gain channel	Lower-gain channel
Input impedance	250 kΩ	250 kΩ
Bandwidth; +0, -3 db	0.3 to 150 Hz	0.3 to 150 Hz
Full-scale sensitivity	900 μv, p.p.	3000 μv, p.p.
Noise level	10 μv, p.p.	30 μv, p.p.

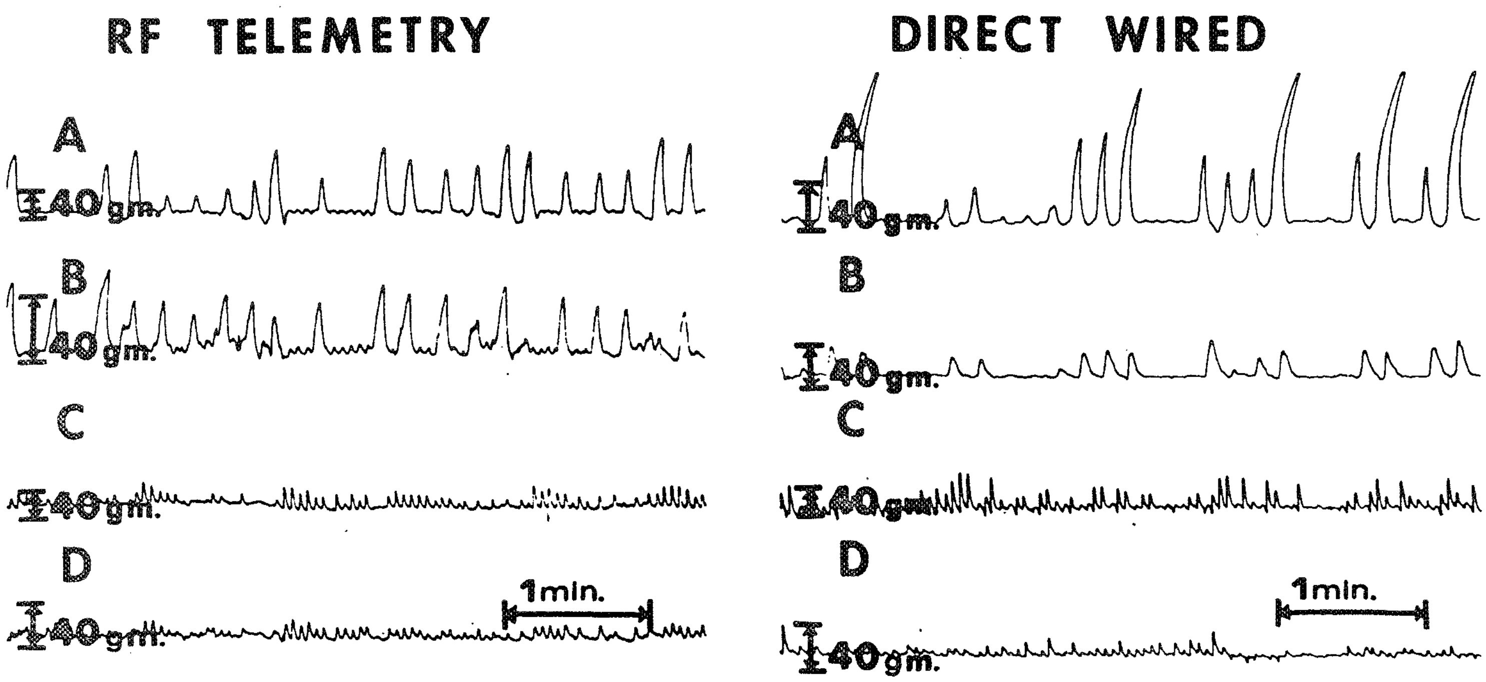
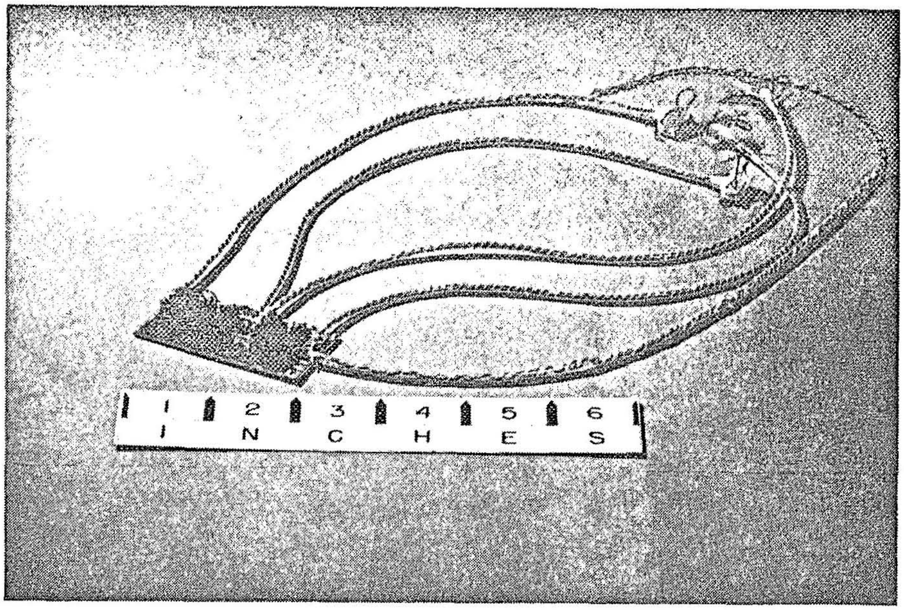
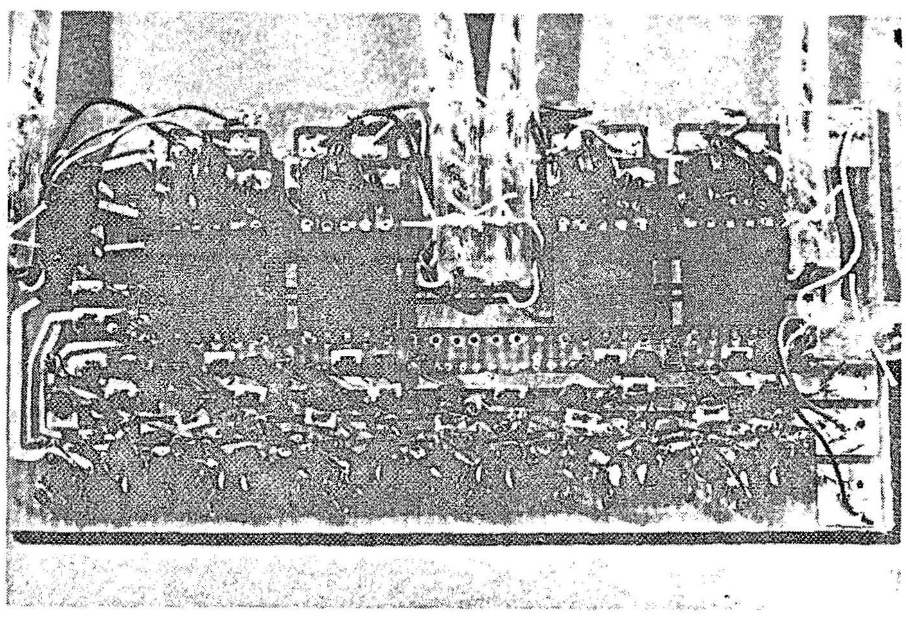


Figure 8.1 Gastro-intestinal contractite-activity of dog. Transducer locations are:

A) Gastric antrum - circular axis
B) Gastric antrum - longitudinal axis
C) Duodenum - circular axis
D) Duodenum - longitudinal axis



a.



b.

Figure 8. 2 Discrete component transmitter.

a) Transmitter with transducers.

b) Closeup of transmitter.

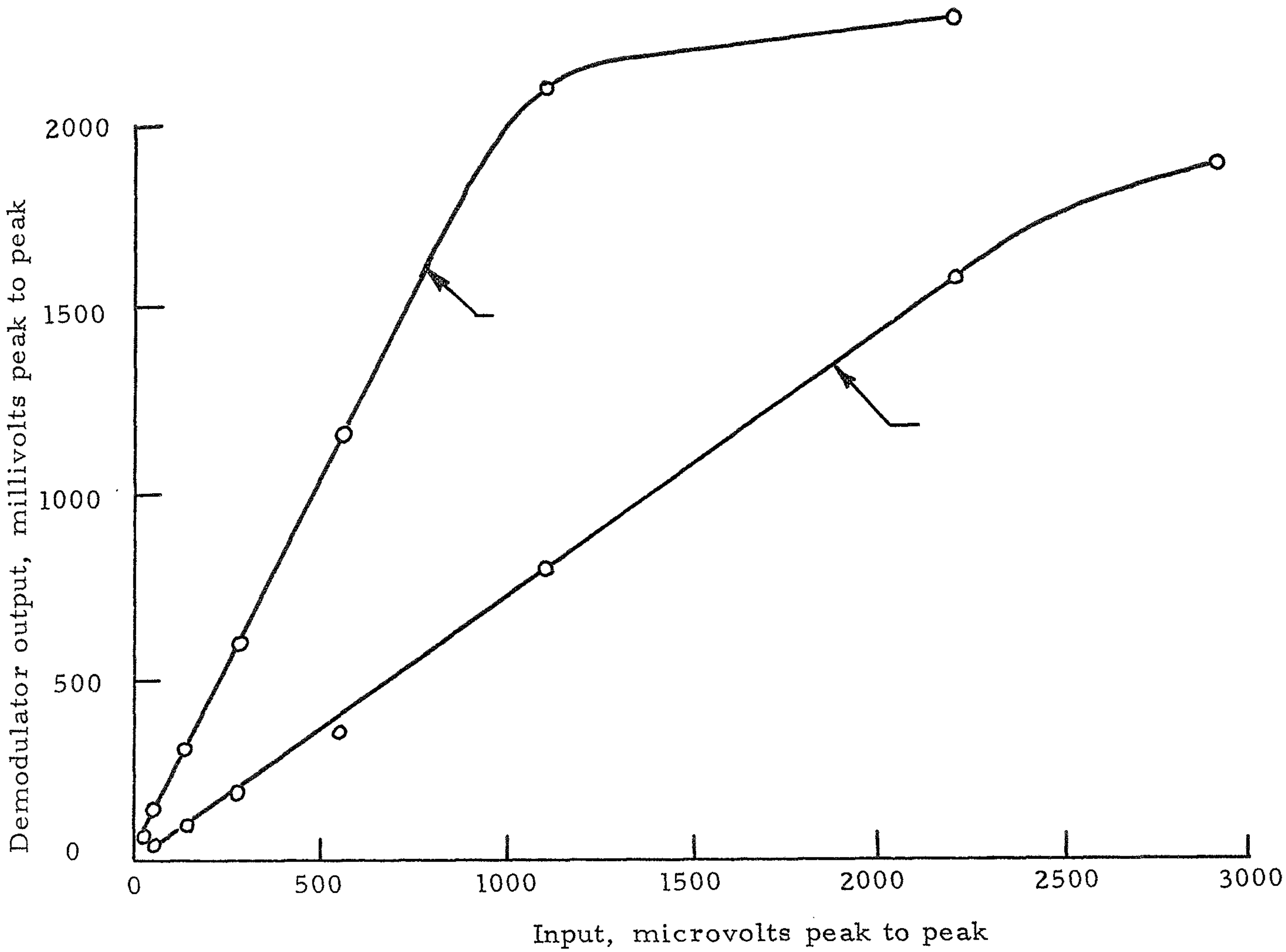


Figure 8.3 Calibration of electrical-signal channels

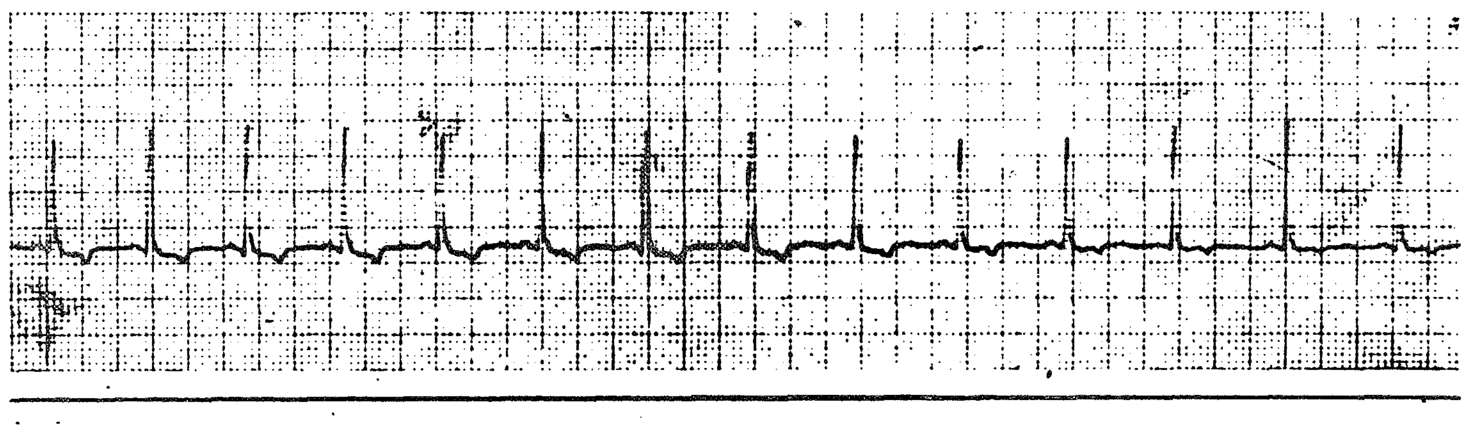


Figure 8.4 Telemetered EKG. Signal picked up by surface electrodes.

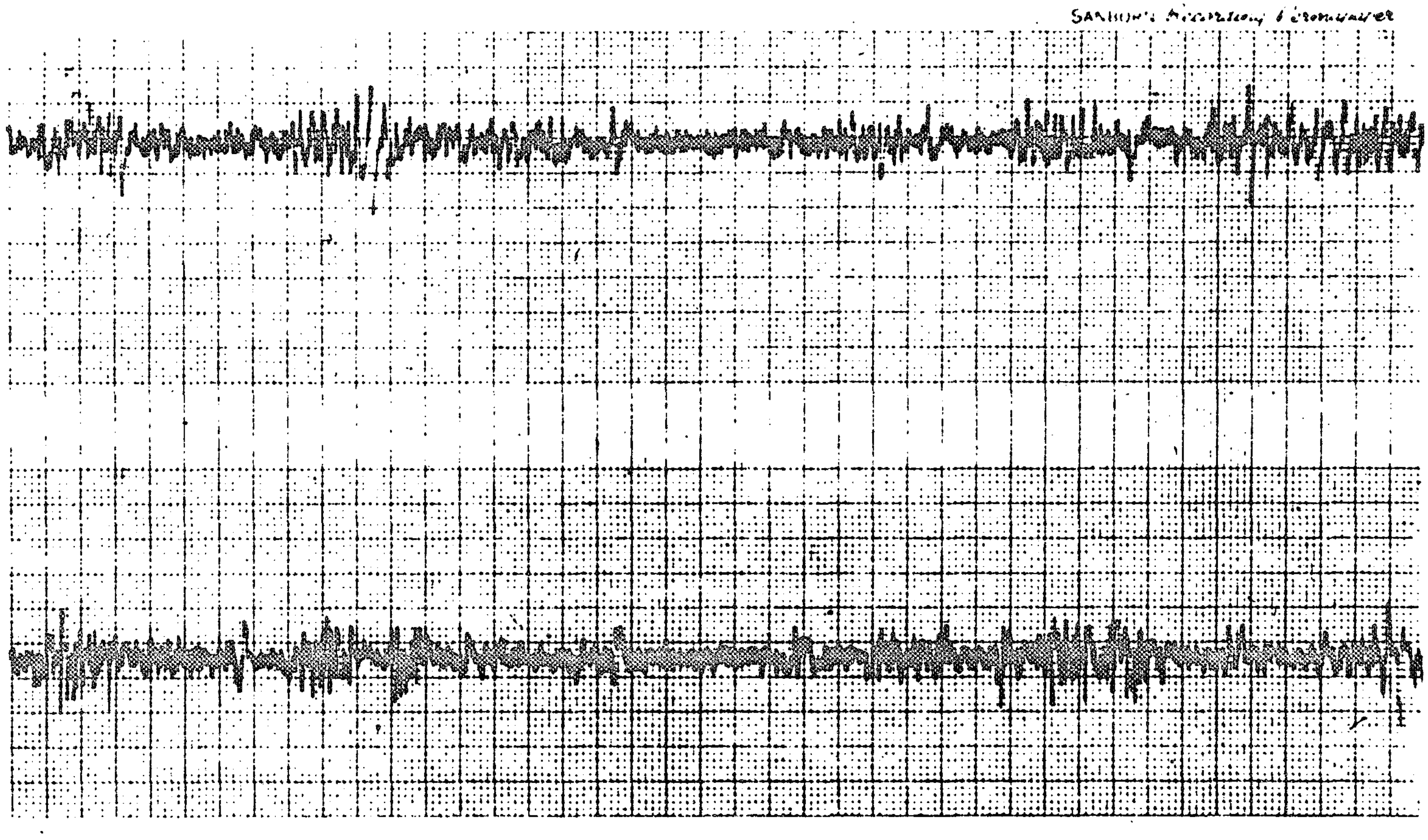


Figure 8. 5 Telemetered EEG. Signals picked up by implanted electrodes in the brain of a cat.

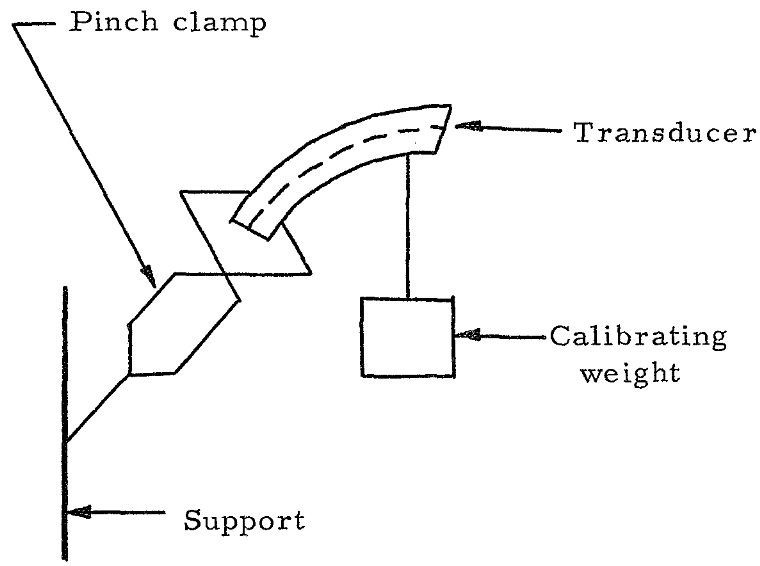


Figure 8.6 Calibrating method for strain-gauge transducers

Force is applied at right angles to center-line of transducers at point of suturing.

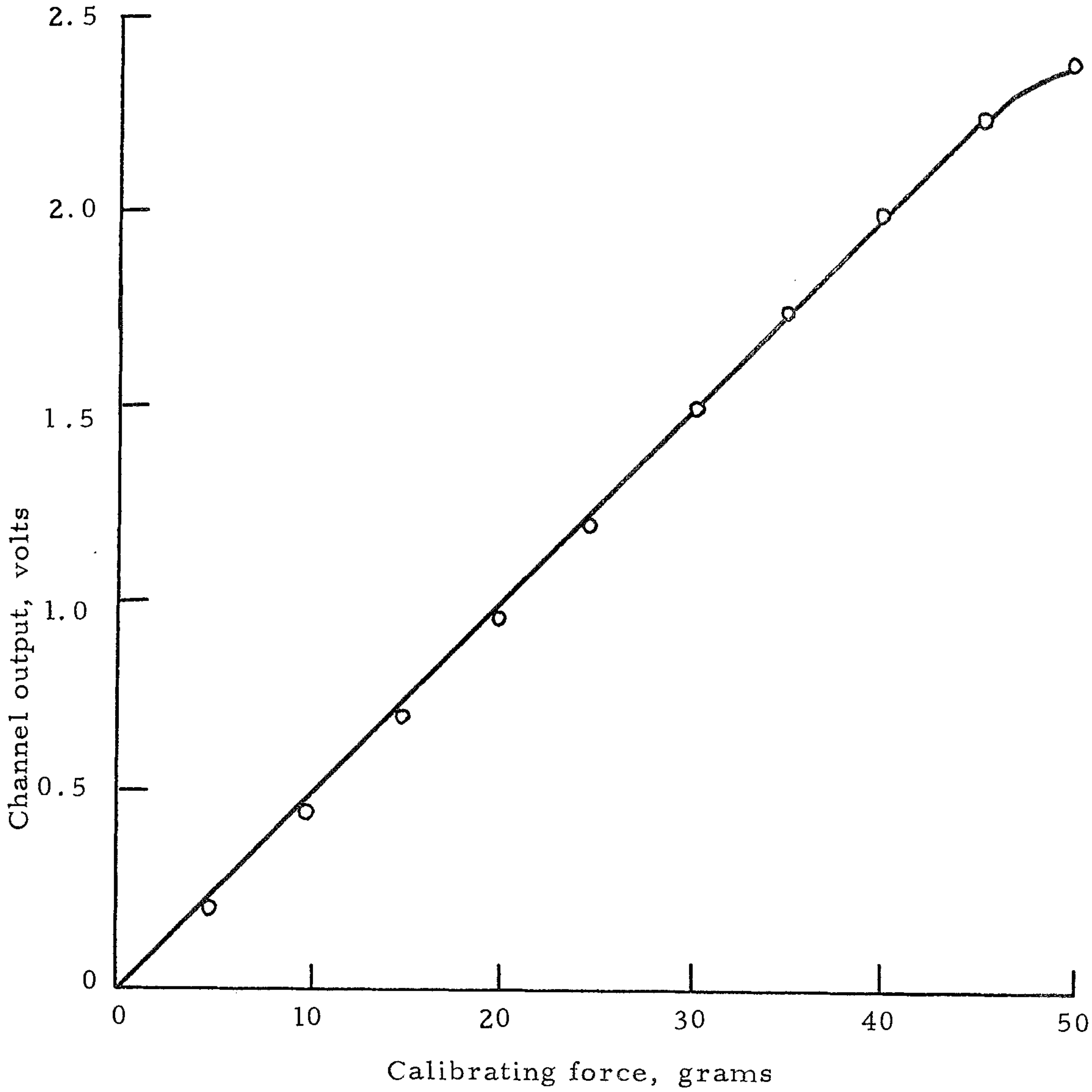
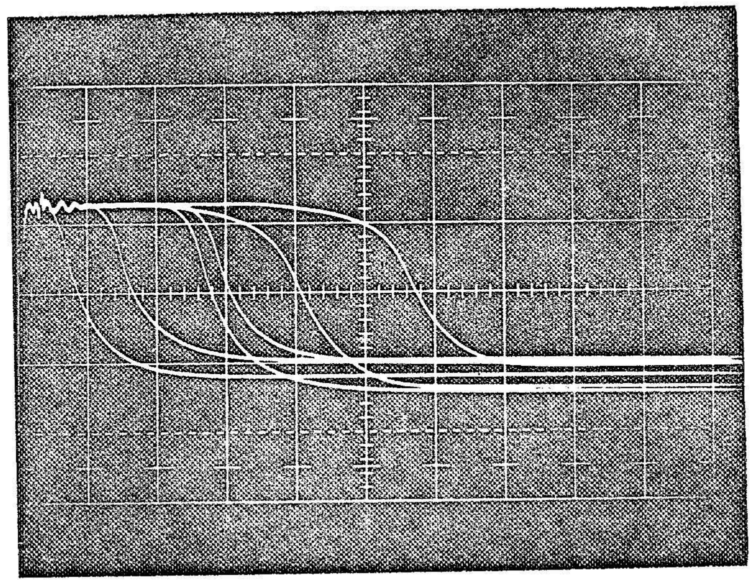


Figure 8.7 Calibration of strain-gauge channel

2 volts per cm



1 microsecond per cm

Figure 8. 8 Scope traces of overlap time variation

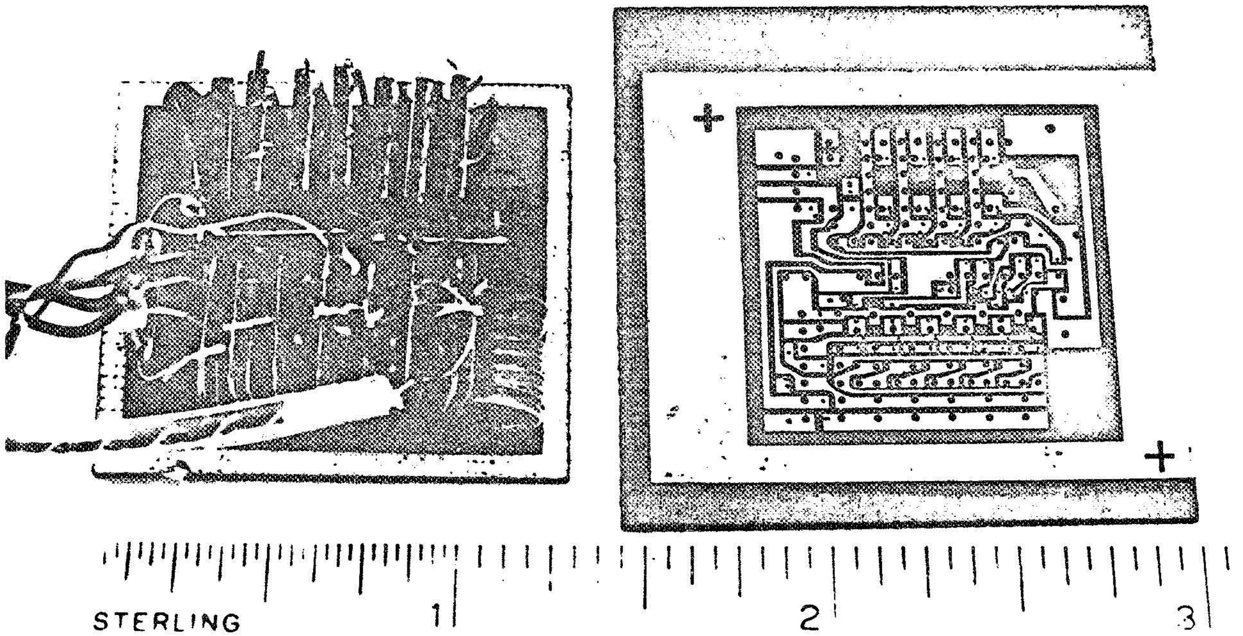


Figure 8.9 Four channel I.C. Transmitter using original circuitry

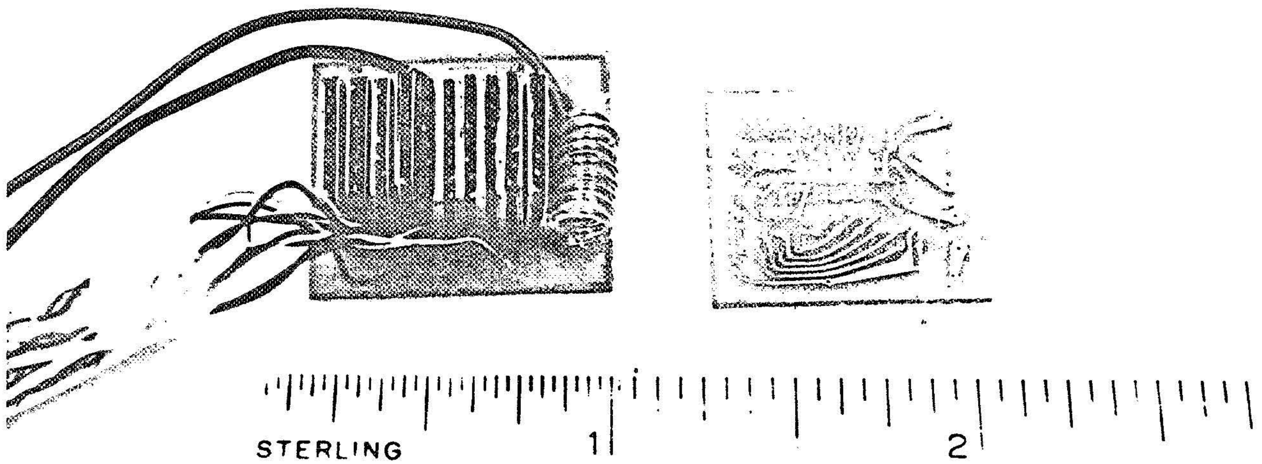


Figure 8.10 Four channel I. C. transmitter using revised circuitry.

SECTION IX

LIMITATIONS AND SUGGESTIONS FOR FURTHER DEVELOPMENT

Transmitter circuitry's switching-times limit the maximum information-bandwidth the system can telemeter. Receiving and demodulating equipment place limitations on the bandwidths available.

Extensions of the system design are possible within the basic framework and circuitry developed. The use of low-duty-cycle operation can be extended to further reduce power-supply requirements. RF-rechargeable battery packs are possible to greatly extend the useful life of the transmitter. Bandwidths can be extended by the use of an AM type RF link.

9.1 Limitations due to transmitter circuitry

Bandwidth is proportional to sampling rate, and sampling rate is limited by switching times of transmitter circuitry. The ring oscillator controls sampling rates. Since the shifting trigger pulse is between 5 and 10 microseconds in duration, and the minimum frame width for each ring oscillator stage is approximately 20 microseconds, the ring oscillator driver frequency upper limit is approximately 40 KHz. This upper limit could possibly be raised by using SCS's with faster turn-on and turn-off characteristics. Waveforms showing the trigger pulse and ring oscillator stage outputs are given in Figure 6.12

9.2 Limitations due to receiver and demodulator

Receiver choice determines RF bandwidth and limits available video bandwidth. FM stereo tuners have RF bandwidths of around 200 kHz and video bandwidths of about 55 kHZ. This will accommodate a 20 kHz sampling rate and 5 kHz of information bandwidth. An FM telemetry receiver with 1 MHz RF bandwidth will have a maximum video-bandwidth of 350 to 500 kHz. This will permit sampling at 100 kHz and give around 25 kHz information bandwidth in an eight channel system.

The revised demodulator circuitry sample window capability of one microsecond is compatible with the maximum sampling capabilities of the present system.

9.3 Extensions of the system

In addition to the method of subcommutation, power supply economy can be realized, at the expense of bandwidth, by low-duty-cycle operation of the entire transmitter. A single frame of samples is taken at a rate within the frame much higher than necessary for information-bandwidth requirements. Then the transmitter is turned off for a period, waiting for the next frame. The frames must have a repetition rate, frame by frame, to satisfy the sampling requirements. This operation could reduce power requirements by a factor of 10 or more in a system with a total-information-bandwidth of 1000 Hz. Receiving and demodulating equipment would require some modification to handle this duty-cycling, especially when the RF carrier is duty-cycled.

Rechargeable batteries would extend the useful life of the transmitter by a factor of 10^3 and reduce the volume of the battery pack to about one cubic inch. Nickel-cadmium batteries can be recharged by coupling an external RF field to an internal pickup coil and rectifying the received RF power[10].

9.4 RF Link improvement

As mentioned previously, the FM RF link may not be optimum for this type of telemetry system. A thorough investigation of the possible RF link alternatives should be made on the basis of D. C. power drain for the minimum usable signal to noise ratio.

Methods of improving the radiation pattern from an implanted transmitter should also be studied to determine an optimum method of coupling power from the implanted transmitter to the receiving antenna.

APPENDIX I

MULTIPLE-CHANNEL PHYSIOLOGICAL TELEMETRY SYSTEMS

Physiological telemetry systems presently available are too large in size and require too much power to be implanted. A survey of typical systems is given in Table A1.1. Power-supply requirements range from 60 to 340 milliwatts. Volumes, including battery packs, are 5 cubic inches and upwards.

Frequency-division multiplexing is used in several designs as it is the easiest to instrument. The subcarrier oscillators consume large amounts of power. Pulse-modulation methods are used in time-division systems. Pulse-duration and pulse-position modulation have been used. In general, these systems require less power than frequency-division systems performing the same telemetry function.

TABLE A1.1

MULTIPLE-CHANNEL, PHYSIOLOGICAL, TELEMETRY SYSTEMS

Modulation Method	Number of Channels	Total Information bandwidth	Power Dissipation	Transmitter Size	Reference
FM/AM	4	400 Hz	340 mw	23 cu. in.	11
FM/FM	6	3200 Hz	60 mw	3.7 cu. in. less batteries	12
FM/FM	4	1600 Hz	340 mw	1.5 cu. in. less batteries	13
AM/FM	10	7500 Hz	not given	not given	14
PDM/FM	7	700 Hz	60 mw	7.2 cu. in.	15
PDM/FM	2 to 6	240 to 720 Hz	not given	4.5 cu. in., approx.	16
PDM/FM	4	8000 Hz	75 mw	6 cu. in., approx.	17
PPM/FM	8	800 Hz	not given	6 cu. in., approx.	18

APPENDIX II

A SURVEY OF TYPICAL PHYSIOLOGICAL SIGNAL PARAMETERS AND MEASUREMENTS

Physiological signals are detected by five basic kinds of sensors:

a) Electrical

b) Force

c) Thermal

d) Flow

e) Impedance

Electrical-signal sensors are either surface or implanted electrodes. Materials used are stainless steel, platinum, silver-silver chloride, and glass with salt bridge. Force sensors are variable impedance elements coupled mechanically or hydraulically to the activity being measured. Temperatures are sensed by thermistors (temperature-variable resistors).

Blood flow measurements are made by electromagnetic and electrosonic flowmeters. Impedance bridges are used to monitor changes in muscle and organ impedances resulting from their activities. Table A2. 1 gives a listing of typical physiological signals with characteristics and measurement methods.

Signals vary in frequency content with direct-electrical signals requiring the highest bandwidths. The remaining signals require bandwidths of DC to 20 Hz. Clinical studies of EMG require 500 Hz bandwidth; EEG and EKG require about 200 Hz. Some studies utilize bandwidths for EEG and EMG to 5000 Hz.

Electrical signals vary in amplitude from 10 to 75 microvolts for EEG to about 4 millivolts for EKG. Nerve and cell-membrane potentials are in the order of 150 millivolts.

TABLE A2.1

CHARACTERISTICS OF PHYSIOLOGICAL SIGNALS

Physiological Signal	Amplitude Range	Frequency Range	Transducer	Reference
Electro-cardiogram (EKG)	0.75-4mv. p.p.	0.1-100 Hz	Electrodes	19
Phono-cardiogram		30-100 Hz	Piezoelectric pickup, microphones	20, 21
Electro-myogram (EMG)	0.1-4 mv. p.p.	$2\text{-}10^5$ Hz 10-500 Hz (clinical)	Electrodes	19
Electroen-cephalogram (EEG)	10-75μv	0.5-200 Hz	External scalp and implanted electrodes	19
Electrogas-trograph	10μv-350 μv	0.05-0.2Hz	Surface electrodes	22
Nerve Potentials	140 mv peak	to 100 pulses/second rise time 0.3 msec. 100/sec. common	Electrode	23, 24
Blood Pressure	0-400 mm Hg	0.5-100 Hz	Strain gauge on artery Hydraulic coupling to transducer	19

Physiological Signal	Amplitude Range	Frequency Range	Transducer	Reference
Blood Flow	1-300 cm/ sec.	1-20 Hz	Electromagnetic flowmeter Ultrasonic flowmeter	26 27
Gastro-intestinal Pressure	20-100 cm H_2O	0-10 Hz	Variable inductance	28
Bladder Pressure	to 100 cm H_2O	0-10 Hz		29
Temperature	90^0 to 110^o F.	0-0.1 Hz	Thermistor Thermal expansion	19
Respiration Rate		0.15-6 Hz	Electrode impedance; piezoelectric devices; pneumograph	19
Tidal Volume	50-1000 ml per breath	0.15-6 Hz	Impedance pneumograph	19
Stomach pH	3-13	0-1 cpm	Glass electrode Antimony electrodes	30
Gastro-intestinal Forces	5-200 grams	0-1 Hz	Strain gauges	9

APPENDIX III

TRANSMITTER PACKAGING

The transmitters must be wired, waterproofed, coated with a physiologically inert material, and sterilized in preparation for implant.

A3. 1. Discrete-Component Transmitter

A 4-channel transmitter was built using subminiature discrete components. Figure 8. 7 gives two views of the strain-gauge channels, multiplexing gates and ring oscillator. This circuitry was fabricated on a printed circuit board and used the following components:

a) NPN transistors - GE microtab, D26E-6

b) PNP transistors - Amperex LDA 451 with 5 mil hard gold leads spot-welded on.

c) 30 milliwatt resistors from British Radio Electronics, Incorporated.

d) Ceramic chip capacitors - Westcap, 5 mil hard gold leads spot-welded on.

A3. 2. Printed-Circuit Fabrication

Care must be exercised in the preparation of printed circuit boards used in the assembly of a hybrid integrated circuit transmitter. Standard printed circuit board techniques of photo resist and etching may be used if extreme cleanliness is observed.

The original artwork was done at ten times finished size, and the reduction was made onto Kodalith Ortho Type 3 film.

Following photo resisting and etching the board was treated with a tin plating solution to prevent oxidation and improve solderability.

Soldering was done with a controlled heat small tip soldering iron and standard 60-40 tin lead rosin core solder.

A3. 3. Integrated-Circuit Fabrication

The component chips are bonded onto a ceramic substrate which has a gold pattern on its surface. Gold paste, containing a glass frit, (Electro-Science Labs, Inc., type 8800B) is silk-screened onto the substrate and fired in at 900°C. The paste has a very viscous vehicle with a low vapor pressure so that the substrates must be inserted very slowly into the firing oven.

The chips are eutectic-bonded, using about 380°C and pressure, onto the patterns. One mil gold wire is thermocompression bonded to the chips to interconnect them. Figure A3. 1 shows a flatpack containing a wired circuit and one with a cover sealed onto it.

The circuits are tested before the lid is put on and corrections are made if necessary. After sealing, the circuit is tested and color-coded to identify it.

A3.4. Flatpack Interconnections

Flatpack integrated circuits are interconnected using a printed circuit board. The flatpacks are inserted on edge. Figure 8.10 shows the printed-circuit board for a 4-channel transmitter, using strain-gage channels, and a completed transmitter. This board has provision for additional discrete components which are necessary to set up the modulation level and to balance the strain gauge signal amplifiers.

A3.5. Packaging and Waterproofing of the Discrete Component Transmitter

The completed circuitry is potted in clear epoxy (Hysol resin, RA2038, and hardener, H2-3404). All lead wires coming through the surface of the epoxy (input and power-supply leads) are degreased with acetone so that epoxy will seal to them.

The battery pack is potted similarly. The two packs are butted together with a teflon mold-release on the adjacent surfaces. A thin layer of epoxy then is used to cover the input and power supply leads running along the surface. This also serves to bond the two packs together but allows them to be separated in order to replace the battery pack. Figure A3.2 shows a completed package with four strain-gauge transducers.

The epoxy is coated with an adhesive (Dow Corning Type A medical adhesive). Nylon mesh is then wrapped around the epoxy package and tied in place with silk sutures, (No. 0). The ends of the sutures are about ten inches in length and extend through the outer silastic covering. There are eight pairs of sutures and these are used to anchor the transmitter inside the animal. The mesh adds strength to the silastic and allows for sutures to be placed at other locations on the transmitter if necessary. This is shown in Figure A3.3.

The final covering (Dow Corning, No. 382, Medical Grade Elastomer) is applied over the entire package.

A3.6. Sterilization

The transmitter is sterilized in Zephiran chloride for 48 hours. A 48-hour water soak follows to get rid of the Zephiran chloride. Distilled, and boiled, water is used for the rinse, with two or three complete changes of water. The unit is then stored in a sterile container, filled with sterile water, until being implanted.

A3.7. Packaging of Other Transmitters

Since complete epoxy encapsulation of the transmitter components makes it almost impossible to remove a component without total destruction of it and surrounding components. It is hoped that a method of encapsulation will be found which will allow removal and testing of defective components to assist in later development and circuit improvement.

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