PERSONAL TELEMETRY TRANSMITTER SYSTEM

JAMES A. ALMOND

FOREWORD

The research and development described in this report was initiated by the Biophysics Laboratory of Aerospace Medical Research Laboratories, Aerospace Medical Division, Air Force Systems Command, Wright-Patterson Air Force Base, Ohio. The research and development was performed between January 1964 and March 1965 by James A. Almond, Design Engineer of Repco Incorporated, Orlando, Florida, assisted by C. M. Williams, Jr., under Contract No. AF 33(615)-1194, Project No. 7222 "Biophysics of Flight." Dr. Adolf R. Marko was contract monitor for the Biophysics Laboratory.

This technical report has been reviewed and is approved.

J. W. Heim, PhD
Technical Director
Biophysics Laboratory

ABSTRACT

The research and development described in this report resulted in the redesign and miniaturization of a Personal Telemetry Transmitter System originally developed by AMRL. The Personal Telemetry Transmitter System transmits seven channels of physiological data on the commercial FM band to a receiver located up to 200 feet away. The seven channels transmitted by pulse duration modulation are electroencephalogram, two leads of electrocardiogram, galvanic skin resistance (base resistance and specific response), respiration, and body temperature. Extremely compact packaging combined with miniature components resulted in a package size of 4.28 by 2.19 by .81 inches for a total volume of 7.6 cubic inches including the battery.

PERSONAL TELEMETRY TRANSMITTER SYSTEM

SECTION I

INTRODUCTION

A Personal Telemetry System is one which allows collection of data from a subject without wires or other direct connections between the subject and the recording equipment. It consists of data collecting, processing, and transmitting equipment on the subject and receiving, decoding and recording equipment at the receiving station. This report will be concerned only with the equipment placed on the subject, the Personal Telemetry Transmitter System. This equipment represents a redesign and miniaturization of the transmitting portion of a Personal Telemetry System developed and reported[1] by AMRL.

The Personal Telemetry Transmitter System is used to transmit physiological data from a subject to a receiving-recording station up to 200 feet away. Since the equipment is placed on a subject that is free to move about, strict limits on size and weight must be observed to insure his freedom of movement. In addition, the equipment must be capable of performing satisfactorily in any environment to which the subject may be exposed.

The Personal Telemetry Transmitter System transmits seven channels of physiological data on the commercial FM band (88-108 MCS). The seven information channels and a sync gap are time-division multiplexed onto one RF carrier. The seven information channels are sequentially sampled and the amplitude information contained in each channel is converted to pulse duration information. The modulating signal then consists of seven rectangular pulses with the width of each pulse corresponding to the signal amplitude of the respective channel. This signal is used to frequency modulate the RF carrier by shifting the carrier frequency by a fixed amount during each pulse period; hence, a PDM-FM signal. The modulating signal pulse format is shown in Figure 1.

In Figure 2 a block diagram of the Personal Telemetry Transmitter System is given. A unijunction oscillator simultaneously drives a three-bistable counter and a sawtooth generator at 1600 pulses per second. From the counter six outputs are obtained which are combined in diode AND-gates to form eight 3-digit combinations. Seven of these combinations are used to sequentially sample the signal inputs while the eighth is used for a sync gap.

The signal from each channel is added to the sawtooth in a resistive network and this signal is gated to the modulator when all inputs from the counter to that gate go positive. The input to the modulator will then be a sawtooth with a DC or low frequency AC component that is clipped at the firing point of the modulator.

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[1] Marko, A.R., Research and Development on Pulse-Modulated Personal-Telemetry Systems, AMRL-Technical Documentary Report 63-96, 6570th Aerospace Medical Research Laboratories, Wright-Patterson Air Force Base, Ohio, October 1963, AD 425574.

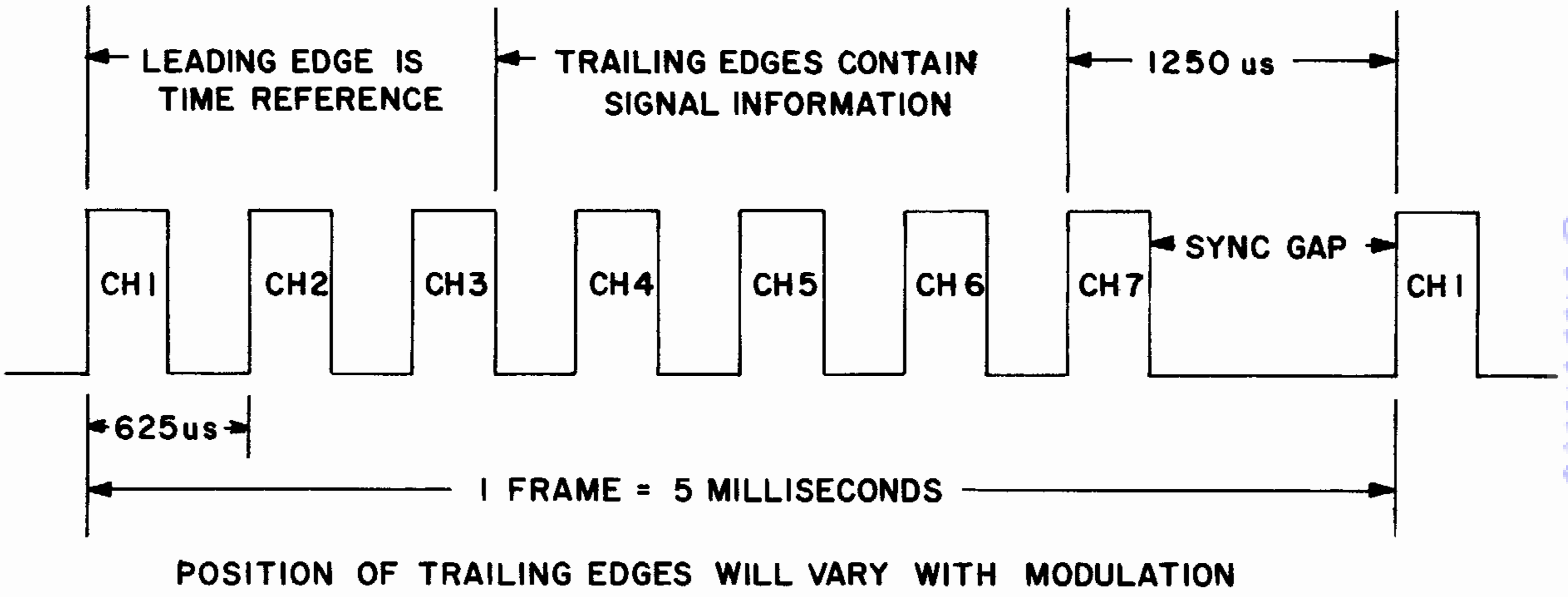


Figure 1. PULSE FORMAT OF SEVEN-CHANNEL PERSONAL TELEMETRY TRANSMITTER

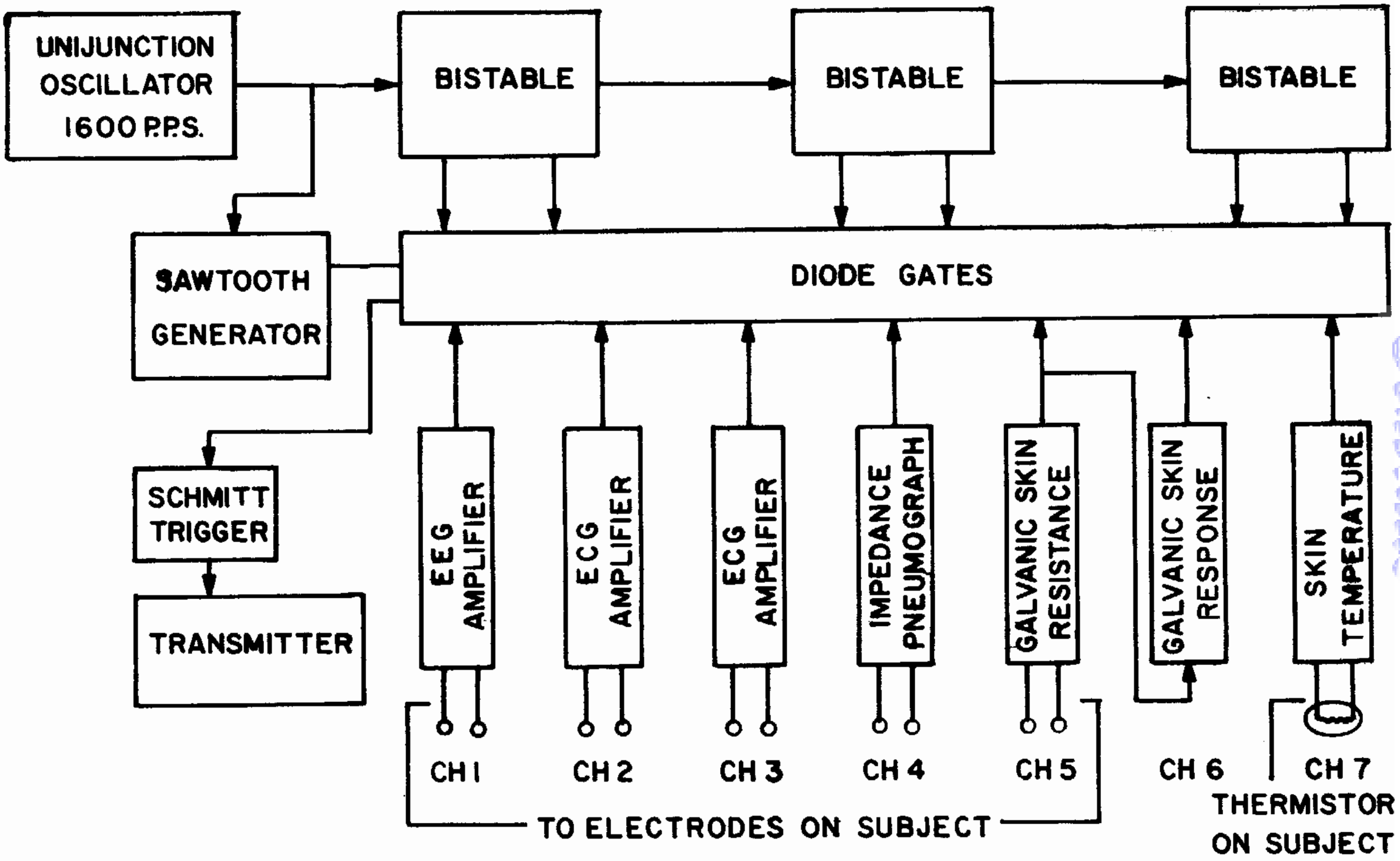


Figure 2. BLOCK DIAGRAM OF SEVEN-CHANNEL PERSONAL TELEMETRY TRANSMITTER

The modulator is a Schmitt trigger having a fixed firing point. The modulator output will be +Vcc during the time that the input exceeds the firing voltage. At all other times the output will be a voltage slightly above ground.

The transmitter is a tuneable oscillator covering the commercial FM band of 88 to 108 mcs. It is frequency modulated by applying the modulator output through a resistor to the oscillator emitter. This method was found to produce adequate deviation while causing very little amplitude modulation.

SECTION II

CIRCUIT DESCRIPTION

Differential Amplifiers - Channels 1, 2, and 3

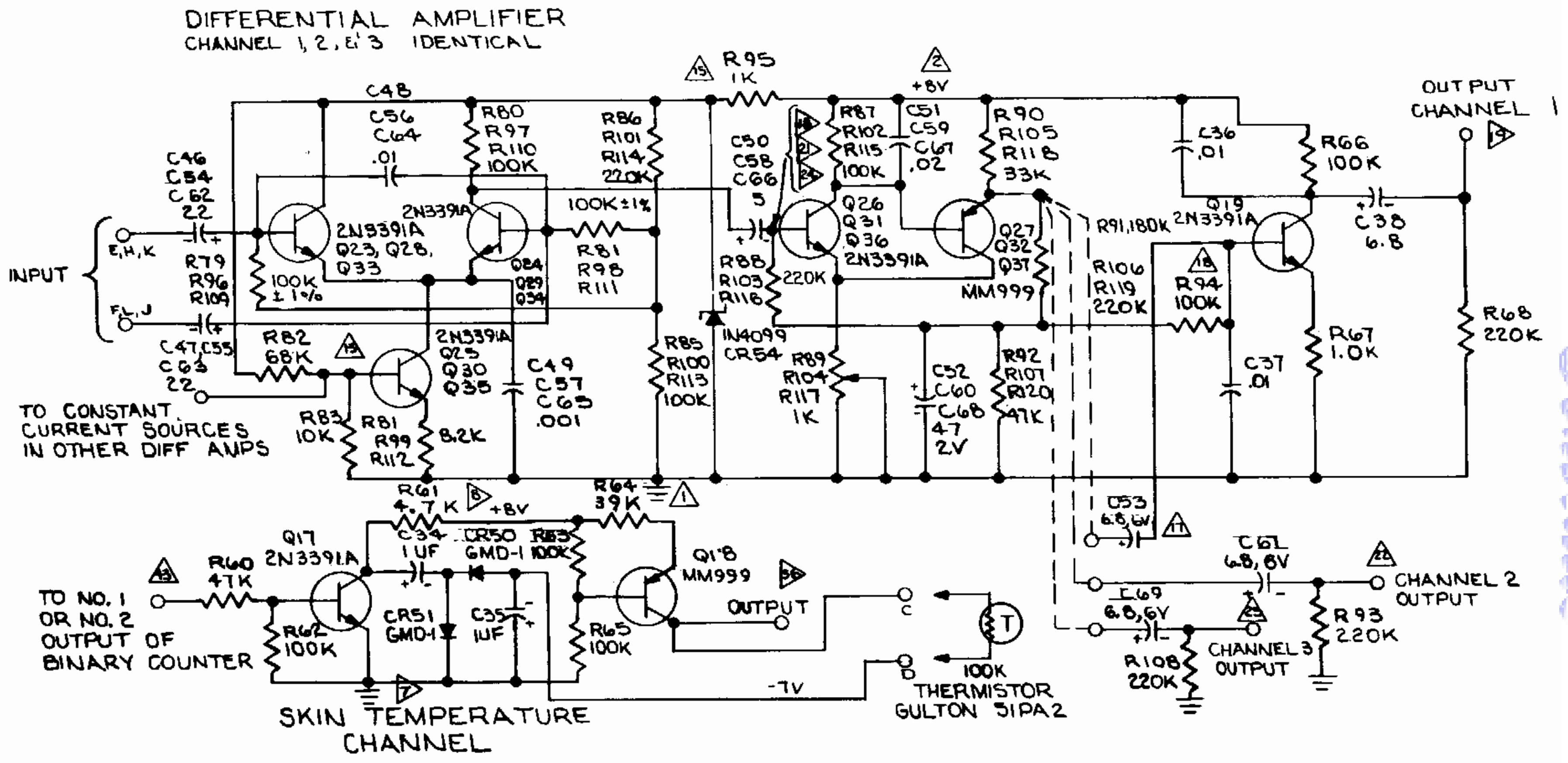
Channels 1, 2, and 3 use high input impedance, differential input amplifiers. The high input impedance is achieved by biasing both transistors from a common bias network through large resistors and by using a constant current source having high impedance as the emitter resistor. The effect is to cause the input impedance to be almost entirely determined by the bias network if low leakage, high beta transistors are used (see Figure 3).

In addition to high input impedance and high gain, the differential amplifier was required to have greater than 60db common mode rejection. Referring to the schematic given in Figure 3, while the input to channels 1, 2, and 3 is differential, the output is taken single-ended. This type operation is necessary since the normal differential amplifier operation (where the output is taken from a transformer with primary coupled between collectors) is impractical for 0.5 to 100 CPS. When a single-ended output is used, common mode rejection can only be obtained by holding the total collector current constant for all conditions of signal input. If a constant current source is used as the emitter resistor, the current division between differential transistors will be determined by the Vbe-Ic characteristics of the transistors and the closeness of Vbe match for the two transistors will greatly affect the common mode rejection.

With the common bias network used for the differential transistors, small differences in base bias resistor values caused a loss of common mode rejection due to the voltage developed by the base current. To minimize this problem, 1% resistors were used for base bias.

Large differences in the value of C46 and C47, such as a +10% variation of one and -10% variation of the other, caused a deterioration of the common mode rejection at low frequencies. A very low frequency common mode signal appeared at the differential transistor bases at a small leading phase angle with respect to the driving signal due to the reactance of C46 and C47. If the values of C46 and C47 were very different, the common mode signal on the base of Q23 was out of phase with that on the base of Q24 causing an excessive common mode output. In general, C46 and C47 were close enough to the nominal value that a low frequency common mode output was not a problem. However, in cases where a problem existed, C46 and C47 were selected to be within 5% of each other.

Radiated RF energy was detected in the base-emitter junctions of the differential transistors causing bias shift and distortion. The bases were coupled together with C48 so any RF signal would appear on both bases and any detected



NOTE:
1. TOP OR FIRST REF NO. IS CHANNEL 1 SECOND CHANNEL 2 & THIRD CHANNEL 3.

Figure 3. SCHEMATIC, DIFFERENTIAL AMPLIFIER AND SKIN TEMPERATURE

voltage would appear as a common mode signal. The emitters were also grounded to RF by C49 to reduce RF pickup by the constant current generator and its wiring.

A single bias network, R82 and R83, was used for all three constant current sources to reduce components and to keep the current equal in the three differential amplifiers.

The channel 1 differential amplifier has approximately 95db voltage gain while channels 2 and 3 each have about 65db. This amount of gain makes it necessary to decouple the differential amplifier stage from the supply voltage to prevent the amplification of any noise or transients appearing on the supply. Since an RC decoupling network would become ineffective at low frequencies, the most simple and effective decoupling is by a resistor-zener diode network such as R95 and CR54. The IN4099 (CR54) was selected for its low current characteristics. Published specifications at 250 uamp Iz, are: zener voltage, 6.8 volts and maximum zener impedance of 200 ohms.

The requirements placed on the amplifier stages following the differential stages are high gain, high input and low output impedance, low frequency response to 0.5 cps, and gain control range of 3 to 1 all at low collector currents. A direct-coupled pair using an NPN common emitter direct coupled to a PNP common collector stage was selected. The first stage bias was derived from the second stage emitter voltage through a dividing and decoupling network. This type of bias along with low collector current in Q26 gives a high input impedance for Q26 which is necessary to prevent loading the output of Q24 and to keep C50 to a reasonable size. The input impedance is maintained at low frequencies by bypassing the degenerative AC feedback with C52. Gain is controlled over a 3 to 1 range by the use of degeneration in the emitter of Q26. The collector current of Q27 is fed through R89 to produce additional degeneration. Through this point channels 1, 2, and 3 are identical.

Channel 1 requires an additional 22db voltage gain above that of channels 2 and 3. This is obtained by the use of an additional common emitter stage in channel 1. Bias for this stage is taken from the same point as that for Q26. RF bypass capacitors are necessary on the emitter and collector of this stage to minimize bias shifting and distortion caused by detection of RF by the transistor junctions.

RC Oscillator

A phase shift RC oscillator running at approximately 15 KCS supplies the driving signal to channels 4, 5, and 6 (Figure 4). The oscillator design is conventional but the output is fed through a small capacitor to back-to-back zener diodes going to ground. This clips the output to a constant level of approximately 4 volts peak-to-peak and prevents changes in drive level with supply voltage change. This signal is then used to drive the signal conditioning circuitry of channels 4, 5, and 6.

Respiration Circuit - Channel 4

The respiration sensor amplifier is a common emitter stage with the base tied to the emitter through a resistor. This causes the transistor to be cut off until drive is applied. The clipped sine wave from the phase-shift oscillator drives the sensor amplifier producing collector current pulses. The amplitude of the collector current pulses will be determined by the driving voltage and the emitter resistor. The emitter resistor is variable to adjust for respiration

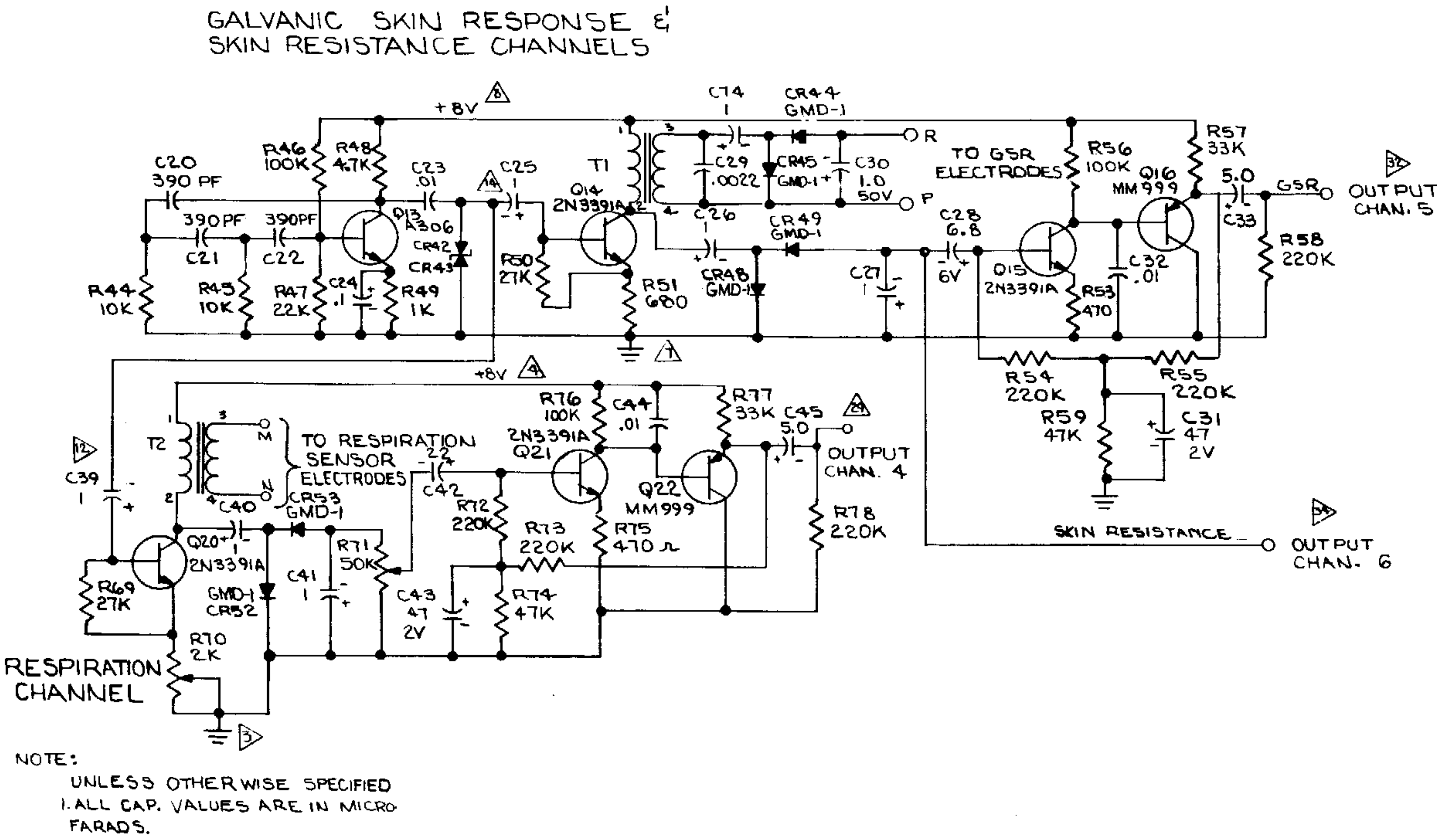


Figure 4. SCHEMATIC, RC OSCILLATOR, GSR AND RESPIRATION CHANNELS

mercury strain guage impedances between 6 and 16 ohms but should never be zero. For emitter resistors greater than a few hundred ohms, the emitter current for a constant amplitude drive will be directly proportional to the emitter resistance; therefore, the collector current will be a constant.

The collector load for the respiration amplifier is a transformer with the secondary going to the respiration mercury strain gauge sensor. The strain gauge impedance is reflected to the transformer primary and, being driven by a constant current, a collector voltage is developed that is proportional to the strain gauge impedance. The collector voltage is rectified in a voltage doubler detector and filtered. A DC output voltage that is proportional to the strain gauge impedance is the result.

Since the depth of respiration will vary from one subject to another, the amount of output can be controlled by R71. The AC component of the detected signal is of interest since it is the change in output caused by respiration. The impedance change of the respiration sensor is less than 2% over the respiration cycle; therefore, the detected output is on the order of 50 millivolts peak-to-peak. Amplification of this signal is effected by use of a direct-coupled common emitter stage and common collector stage identical to that used in the differential amplifier. This circuit combines high gain and high input and low output impedance at very low collector currents.

Galvanic Skin Response and Skin Resistance - Channels 5 and 6

The Galvanic Skin Response (GSR) amplifier is driven by the clipped sine wave supplied by the phase shift oscillator. The transistor base is connected through a resistor to the emitter, keeping the transistor off except when driven, and increasing the input impedance during the time it is driven. For a constant amplitude input, the emitter current will consist of positive going current pulses with the amplitude being determined by the emitter resistor. With a constant collector current, the AC collector voltage will be directly proportional to the impedance in the collector circuit. The collector load is the primary of transformer T1. The secondary of T1 is broadly tuned to 15 KCS by C29. The transformer secondary voltage is rectified in a voltage-doubler detector, filtered, and applied to the GSR electrodes. By using a voltage-doubler detector, the load seen by the transformer secondary is only half that across the GSR electrodes. Therefore, this circuit requires only half the secondary turns that would be required for a full wave bridge detector. The transformer has a 1:2 primary-to-secondary turns ratio so a total impedance step-down of 8:1 from GSR electrodes to transistor collector is effected.

Since the AC collector voltage of the GSR amplifier is directly proportional to the impedance appearing across the GSR electrodes, this voltage will be rectified in a voltage-doubler detector and filtered for the skin resistance (channel 6) output. This will be a negative voltage; it will be most negative for high skin resistance (100k) and approach zero for the lowest skin resistance (5k). The output from channel 6 will at all times be directly proportional to the absolute skin resistance.

As small changes in skin resistance occur, the detected output voltage for channel 6 will vary. These short-term variations in skin resistance are the Galvanic Skin Response. The skin response output is so small that a direct coupled pair, a common emitter stage followed by a common collector stage, is used to amplify the signal to the level required for modulation. This amplifier is similar to the one described in the differential amplifiers.

Skin Temperature - Channel 7

The skin temperature channel circuitry (Figure 3) consists of a negative voltage supply and a constant current source supplying current to a thermistor placed on the subject. The negative voltage is generated by driving an amplifier with one of the outputs from the first bistable in the counter. The collector voltage is a square wave that is approximately 8 volts peak-to-peak. This is detected in a voltage-doubler detector giving a negative supply of approximately 7 volts.

The constant current source supplies current to the thermistor which is connected between the collector and the -7 volts supply. The collector voltage is then a function of the thermistor resistance which is determined by the skin temperature.

Unijunction Oscillator

The center of the timing system is the unijunction oscillator. The oscillator output is a 3 volt negative pulse at 1600 pulses per second repetition rate. In order to maintain good frequency stability, a stable 1% resistor and a Mylar* capacitor are used in the timing circuit. The capacitor, C1, was selected to set the repetition rate to 1600 pulses per second.

Binary Counter

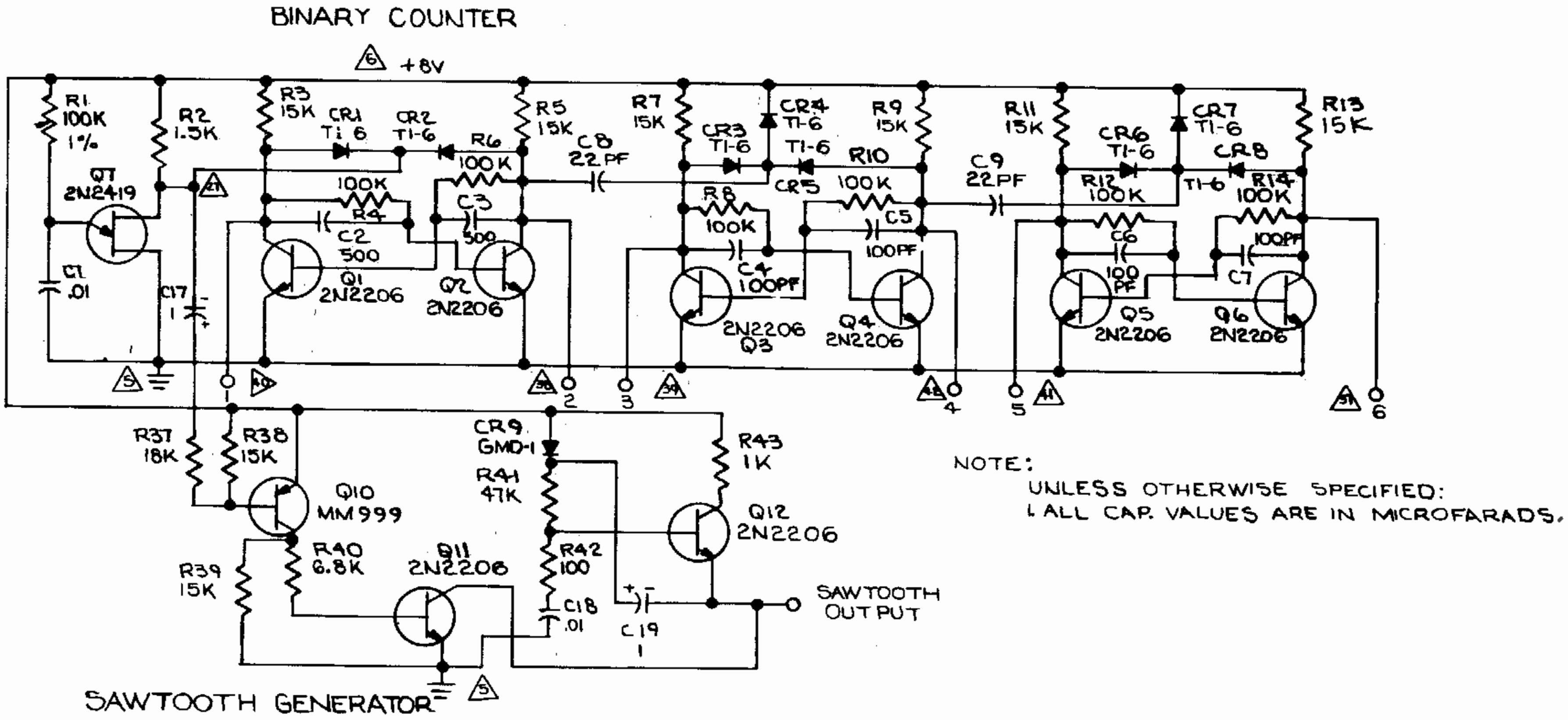
A three bistable counter is driven by the pulses from the unijunction oscillator (see Figure 5). The unijunction oscillator pulse is coupled through steering diodes to the transistor collectors. The pulse is then cross-coupled through "speedup" capacitor C2 or C3 to the opposite base to initiate switching action. In the first bistable, C2 and C3 are large because of the poor rise time characteristics of the pulse from the unijunction oscillator. Large resistors are used in all the bistables to keep current drain to a minimum.

The output of the first bistable is capacitively coupled to the second and the second to third. The second and third bistables are like the first except for smaller "speedup" capacitors and an additional diode. Since the signal driving the second and third bistables is a differentiated square wave having a very good rise time, small capacitors can be used to couple this signal into the correct base. An additional diode is required in the second and third bistable to clip the positive going pulse created when the square wave is differentiated. The outputs from the six bistable collectors are fed directly to the gates.

Sawtooth Generator

The unijunction oscillator output pulse is also used to drive the sawtooth generator which puts out a sawtooth wave that is synchronized with the binary counter output. A bootstrap circuit is used to generate the sawtooth output. The voltage across a capacitor being charged by a constant current will increase linearly. The constant current is supplied by maintaining a constant voltage drop across resistor R41. This is done by increasing the supply voltage at the same rate at which the voltage on the capacitor increases. The voltage across C18 (refer to Figure 5) is amplified by the emitter-follower Q12 and added to the supply voltage through capacitor C19. By proper selection of R41 and C18 the emitter voltage will rise to within 0.5 volts of the supply voltage at the end of the sawtooth.

* DuPont Trademark



NOTE:
UNLESS OTHERWISE SPECIFIED:
1. ALL CAP. VALUES ARE IN MICROFARADS.

Figure 5. SCHEMATIC, BINARY COUNTER AND SAWTOOTH GENERATOR

The sawtooth generator is driven by the unijunction oscillator. The pulse from the unijunction oscillator is amplified by Q10 which drives a switch, Q11. The switch Q11 grounds the emitter of Q12 during the unijunction oscillator pulse, discharging C18 through the 100 ohm resistor and the base-emitter diode of transistor Q12. At the same time C19, the bootstrap capacitor, is recharged to the supply voltage through diode CR9. This replaces any charge lost from C19 during the period of the sawtooth.

The 100 ohm resistor in series with the sawtooth capacitor limits the discharge current through the base-emitter diode to a safe value but has no effect on the sawtooth output. The large base current during the discharge of C18 would cause extremely large collector currents in Q12 if it were connected as a normal emitter follower. This could cause damage to Q12 and would place transients on the supply voltage that would be a constant source of trouble. To eliminate the problem, R43 was added in the collector of Q12 to limit the collector current to a reasonable value during the discharge of C18. The addition of R43 does not appreciably affect the sawtooth output.

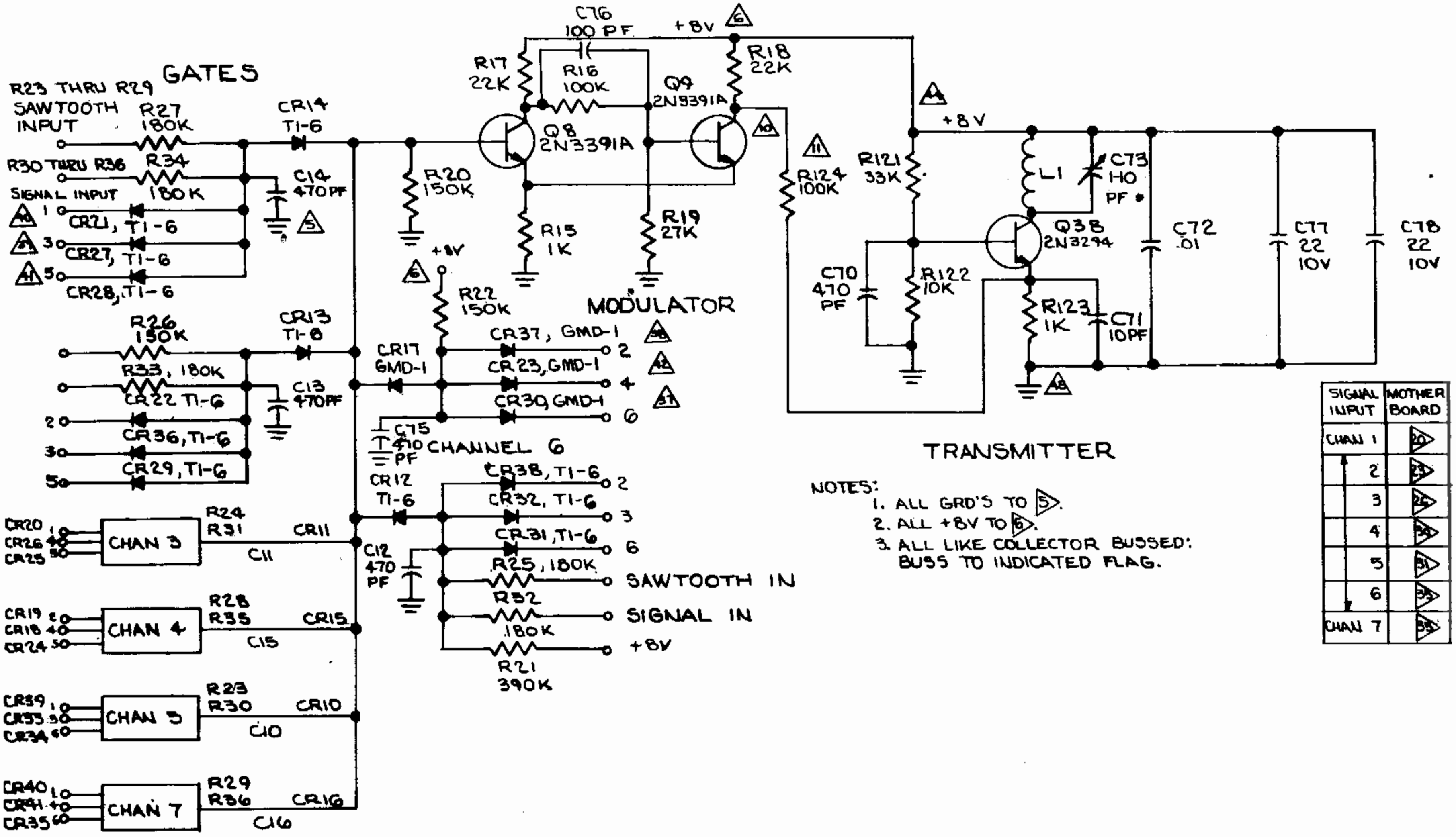
Signal Gates

The seven signal gates are identical in design except for Channel 6. Each gate has a sawtooth input and a signal input which are combined in a resistor network (refer to Figure 6). A three input diode AND gate is connected to the point where the signal and sawtooth are combined. When any input to the AND gate is zero, the signal is shorted to ground and no output is obtained. Once during each frame, or 200 times per second, all inputs to the AND gate will be positive and the signal will be gated through CR14 to the modulator. A small capacitor C14 is added to the signal combination point to bypass any transient produced by different switching times of the bistables or by propagation time of the pulses through the counter. The diode AND gates are driven by the six outputs from the binary counter. The outputs are combined in the AND gates to give eight combinations which occur sequentially and thus time-division multiplex the signals to the modulator.

Channel 6 has an additional resistor R21, which adds +8 volts in with the signal and the sawtooth. This is necessitated by the large negative voltage of the channel 6 signal which is generated by the GSR amplifier. The sync channel also has +8 volts supplied through a resistor R22 to the gate output supplying a positive voltage to the modulator during the sync gap.

Modulator

A Schmitt trigger or squaring circuit is used for the modulator. The signal driving the modulator will be the combination of the channel signals and the sawtooth voltage as gated through diodes CR10-17. This signal will be the sawtooth added to a DC or low frequency AC signal so it will appear as a sawtooth varying in width and amplitude (but not in slope) as the modulating signal varies. An input threshold level is established by the Vbe of Q8 and the Ve developed by the emitter current of Q9. Q8 will be cut off until the threshold level is exceeded at which time Q8 will saturate and Q9 will turn off. A commutating capacitor C76 is used to improve the turn-on and turn-off characteristics of Q9 to give short rise time square waves out of Q9. The output of Q9 will be Ve + Vce (sat) when the input to Q8 is below the threshold and +8V when the input is above the threshold.



NOTES:
1. ALL GRD'S TO 5.
2. ALL +8V TO 6.
3. ALL LIKE COLLECTOR BUSSED: BUSS TO INDICATED FLAG.

SIGNAL INPUT	MOTHER BOARD
CHAN 1	20
2	23
3	25
4	30
5	31
6	33
CHAN 7	35

Figure 6. SCHEMATIC, SIGNAL GATES, MODULATOR AND TRANSMITTER

Transmitter

The transmitter is a Colpitts oscillator having a radiating tank coil, L1. The frequency of oscillation is adjustable with C73 over the range of 88 to 108 megacycles. The transistor operates in the common emitter mode with respect to DC so good DC stability is maintained and as a common base stage to the AC or RF signal. The radiating tank coil, L1, was made as large as practical and uses silver plated wire for highest Q factor and best radiating characteristics. The feedback path for the oscillator is through the transistor collector-emitter capacity and C71. The base is placed at AC ground by C70.

The modulating signal is applied to the emitter. This method of modulation was found to give very little amplitude modulation but very good frequency modulation. The modulating signal causes a small change in emitter current and collector-emitter voltage which produces a change in collector-emitter capacity and shifts the frequency. Approximately 60 KCS frequency shift is obtained with the component values used in this circuit but more or less deviation can be obtained by changing R124, the resistor limiting the modulation applied to the emitter of the transmitter.

Battery

The TR-146 or E146 Mercury battery was chosen as the best compromise between the conflicting requirements of size, life, availability, and connector reliability. The TR-146 is rated at 350 milliampere hours giving better than 30 hours service. It is readily available and the snap connectors offer a reliable connection.

Packaging

The battery, being the largest component, determined the width and height of the case. With these dimensions fixed, the length was set by the volume and the amount of circuitry used. Since most of the components used were not as tall as the box height but were too tall to allow back-to-back printed circuit boards another method was used. A large amount of board area was required to mount components of varying heights. To gain this board area six component boards were used with each of these being wired to and mounted on a motherboard containing interconnecting printed circuitry. Component location drawings are given for each of the boards in Figures 8, 9, and 10. A drawing of the motherboard is shown in Figure 11 with points on the board flagged by numbers in triangles corresponding to points on the schematics given in Figures 3 through 6.

The transmitter and its radiating coil are located on the motherboard on the opposite end from the connector for sensor inputs. The battery with shields on each side lies between the transmitter and the signal processing and multiplexing circuitry to minimize problems caused by RF energy being radiated into this circuitry.

All inputs to the signal processing circuitry are brought out on the 14 pin connector for sensor connections. In addition, pin A goes to the battery negative and pin B to the circuitry ground so power is supplied by connecting A to B on the plug. This also supplies a convenient ground reference for any of the sensor inputs.

The case is made of G-10 epoxy glass cloth board and is cemented with epoxy resin to give a case of exceptional strength and impact resistance. Overall dimensions of the case are 4.28 by 2.19 by .81 inches. Volume of the completed unit is 7.6 cubic inches excluding the connector protrusion. Openings are provided in the rear of the case for frequency adjustment, gain adjustment on each of the differential amplifier channels, and gain and output adjustments on the GSR channels.

SECTION III

CONCLUSIONS AND RECOMMENDATIONS

The equipment described herein was designed and constructed to meet rigid size and weight requirements as well as to perform satisfactorily electrically. Further reductions in size are possible at some increased cost by utilization of integrated circuits. Many integrated circuits have become available during the development of this equipment that could be used with little redesign. Specific areas that could use integrated circuits or multiple transistors or diodes in one package are the bistables in the binary counter, the differential amplifiers, the diode gates, and the direct-coupled pairs of NPN and PNP transistors. Since the battery consumes a large percentage of the volume any reduction of the battery size that would be made without unduly shortening the life would directly reduce the package size.

APPENDIX

COMPONENT AND CONSTRUCTION INFORMATION

T1: PRIMARY 100T NO. 36 NYCLAD
SECONDARY 200T NO. 36 NYCLAD

T2: PRIMARY 300T NO. 36 NYCLAD
SECONDARY 20T NO. 36 NYCLAD

T1 & T2 LAYER WOUND ON MM 472-A FERALEX

L1: TRANSMITTER COIL: 4 3/4 T NO. 20 COPPER

BUS WIRE, SILVER PLATE, AIR CORE, CLOSE SPACED

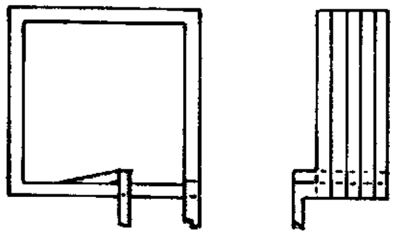
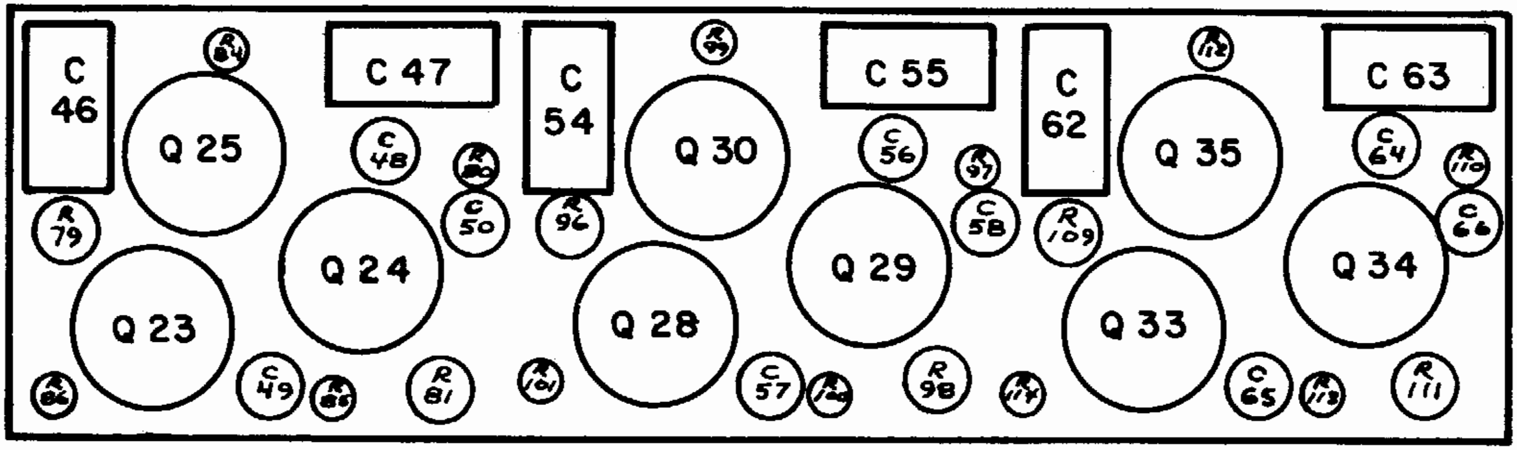
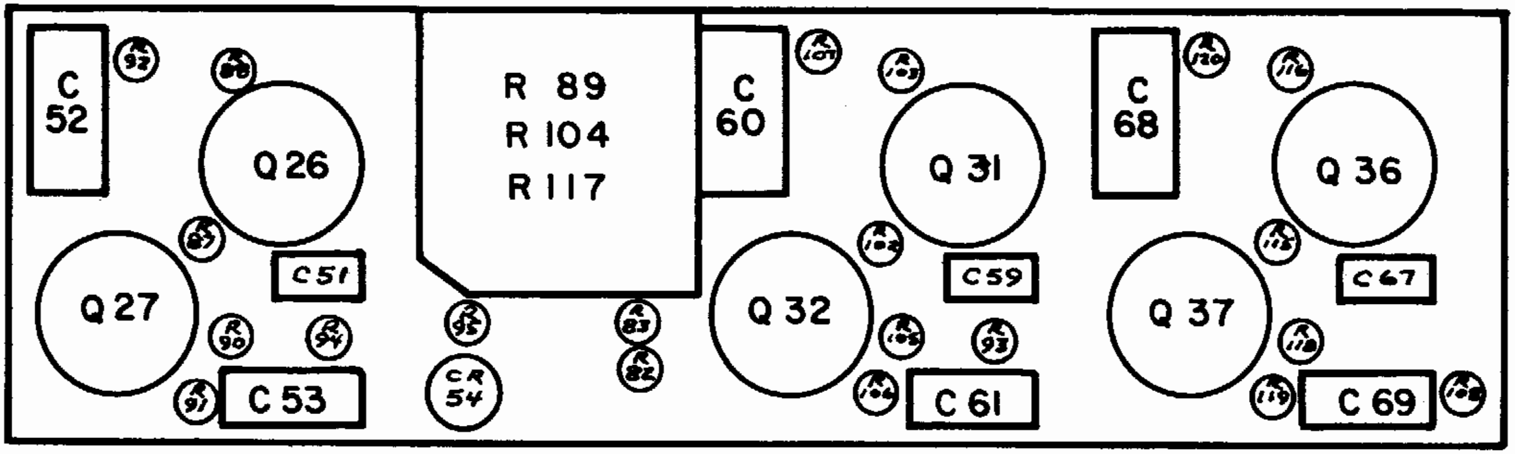


Figure 7.

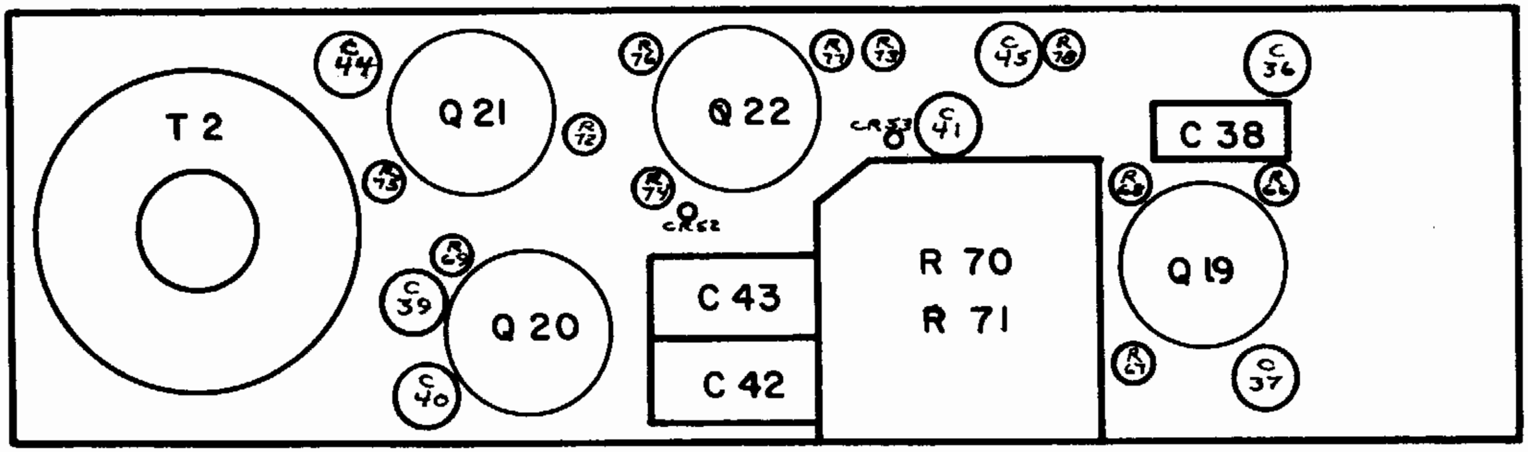


CARD A: DIFFERENTIAL AMPLIFIERS (FIRST THREE STAGES)

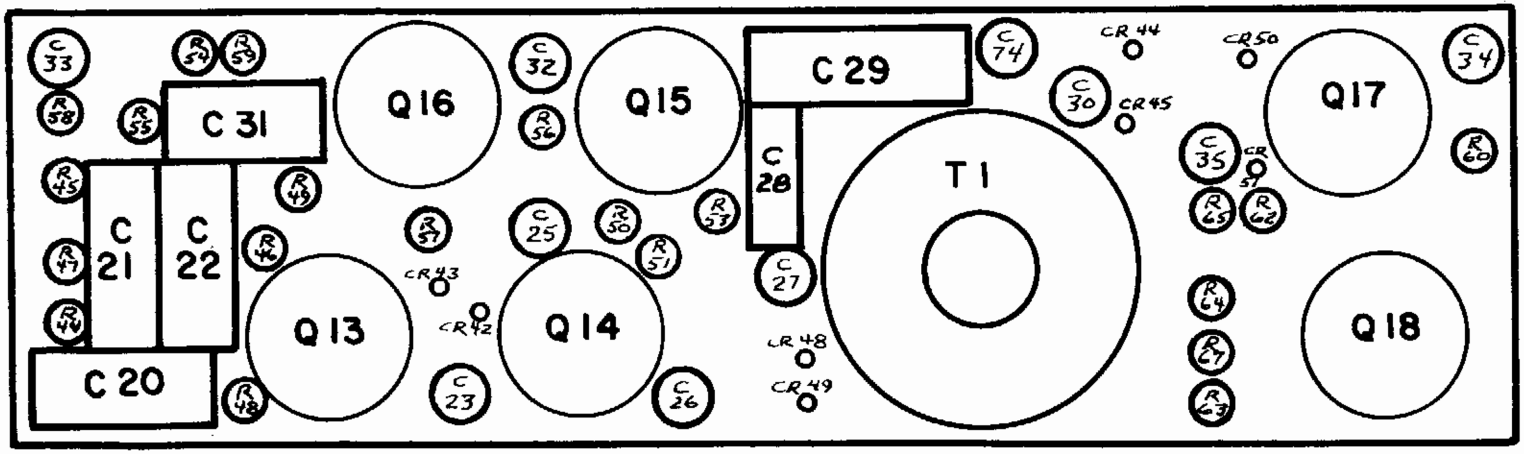


CARD B: DIFFERENTIAL AMPLIFIERS (TWO STAGES)

Figure 8.

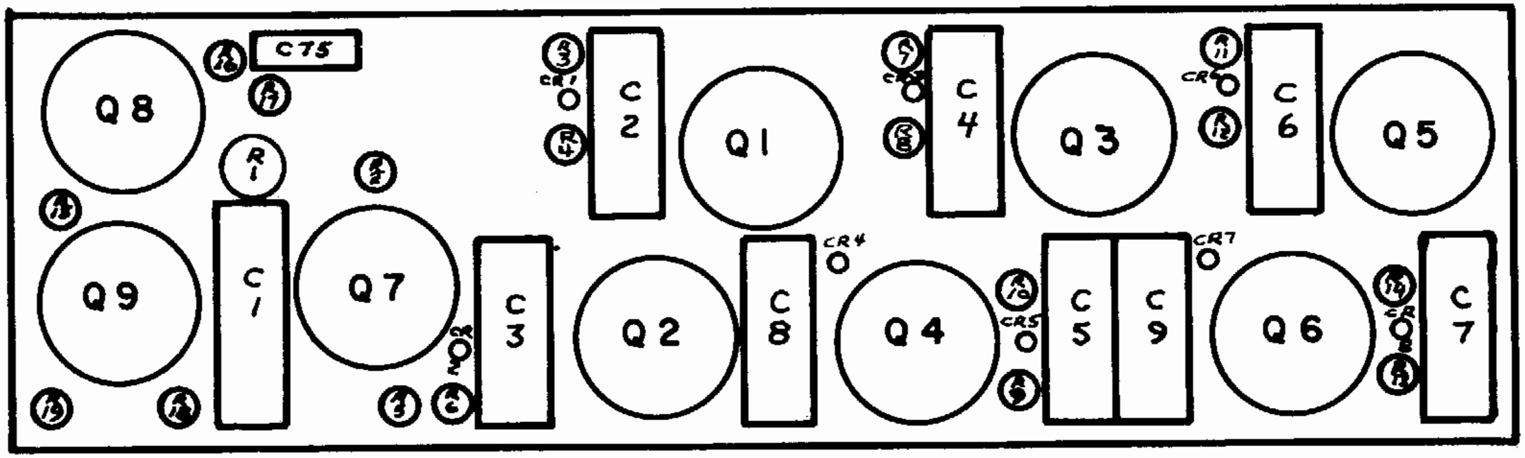


CARD C: SKIN TEMPERATURE CHANNEL & FINAL STAGE EEG

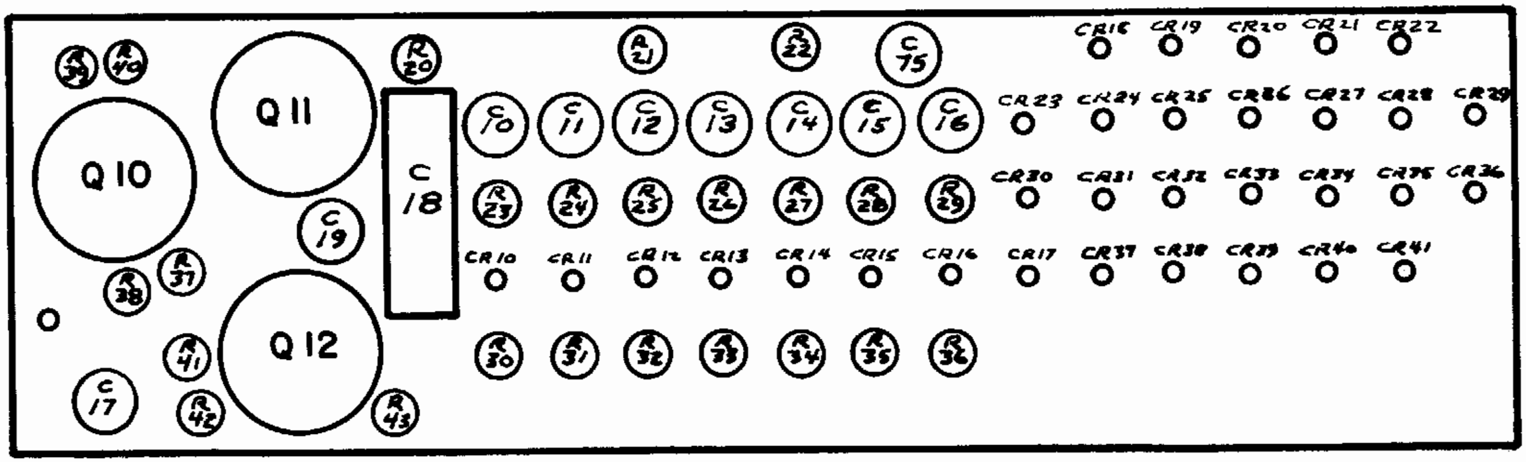


CARD F: GSR & SKIN RESISTANCE CHANNELS

Figure 9.



CARD D: MODULATOR & BINARY COUNTER



CARD E: SAWTOOTH GENERATOR & GATES

Figure 10.

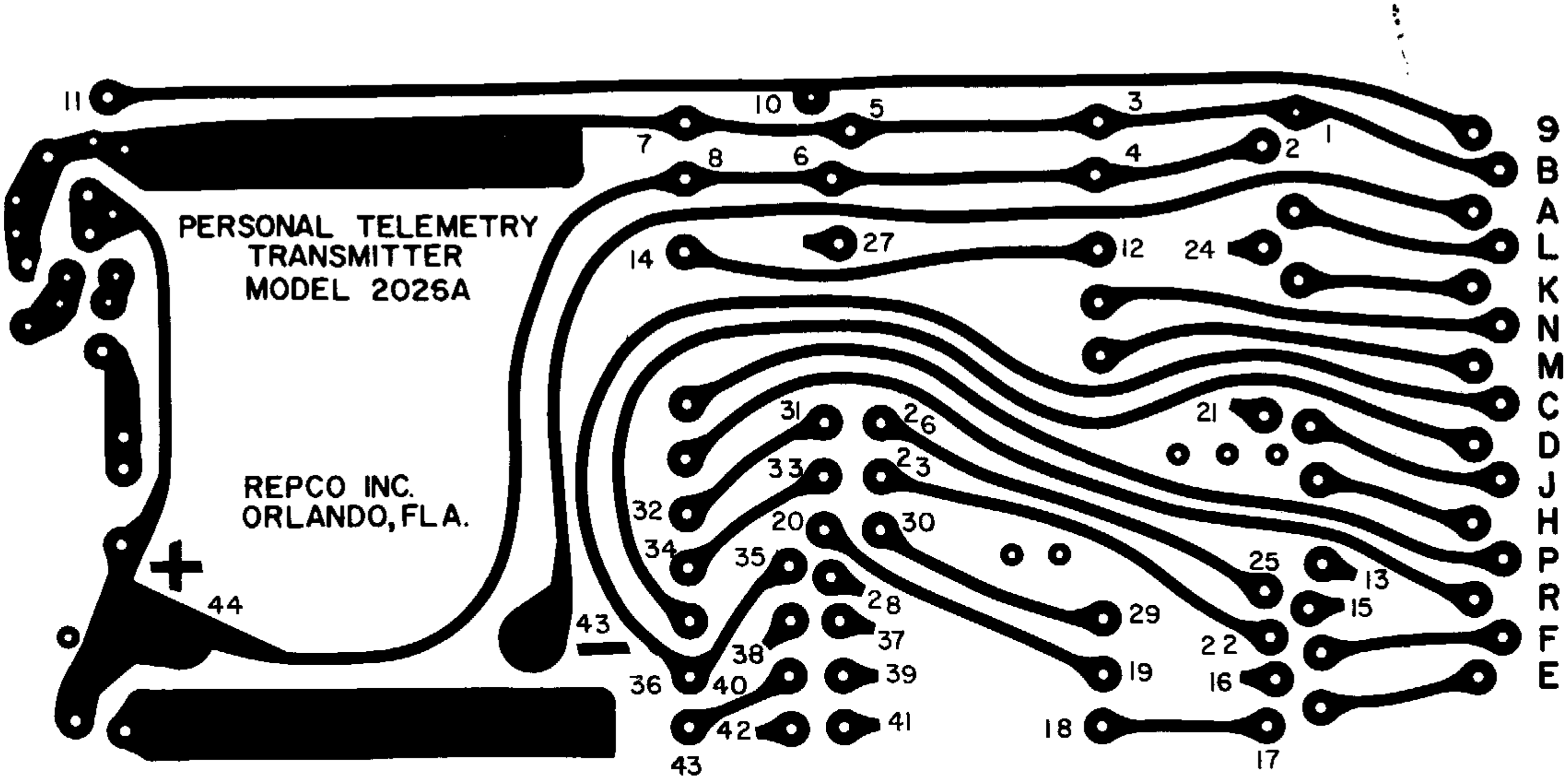


Figure 11. PRINTED CIRCUIT CARD TIE POINTS

PARTS LIST

REF. NO.	PART NO.	DESCRIPTION	REMARKS
C1	4M103	.01 uf Capacitor	Westcap
C2	CD7-501J	500 pf Capacitor	Cornell Dubilier
C3	CD7-501J	500 pf Capacitor	Cornell Dubilier
C4-7	CD6-101	100 pf Capacitor	Cornell Dubilier
C8-9	CD6-220	22 pf Capacitor	Cornell Dubilier
C10-16	390-000X5UD 471M	470 pf Capacitor	Erie
C17	TS01-10-105	1 uf Capacitor	US Semcor
C18	4M103	.01 uf Capacitor	Westcap
C19	TS01-10-105	1 uf Capacitor	US Semcor
C20-22	CD7-391	390 pf Capacitor	Cornell Dubilier
C23	WM12BX103M	.01 uf Capacitor	Westcap
C24	P104R	.1 uf Capacitor	Components
C25-27	TS01-10-105	1 uf Capacitor	US Semcor
C28	M6R8	Capacitor	Components
C29	4M222	.0022 uf Capacitor	Westcap
C30	TS01-10-105	1 uf Capacitor	US Semcor
C31	L47	47 uf, 2v Capacitor	Components
C32	WM12BX103M	.01 uf Capacitor	Westcap
C33	TS02-6-505	5 uf, 6v Capacitor	US Semcor
C34-35	TS01-10-105	1 uf, 10v Capacitor	US Semcor
C36-37	WM12BX103M	.01 uf, 100v, Capacitor	Westcap
C38	M6R8	6.8 uf, 10v Capacitor	Components
C39-41	TS01-10-105	1 uf, 10v Capacitor	US Semcor
C42	L22	22 uf, 10v Capacitor	Components
C43	L47	47 uf, 2v, Capacitor	Components
C44	WM12BX103M	.01 uf Capacitor	Westcap
C45	TS02-6-505	5 uf, 6v Capacitor	US Semcor
C46-47	L22	22 uf, 10v Capacitor	Components
C48-49	WM12BX103M	.01 uf, 100v Capacitor	Westcap
C50	TS02-6-505	5 uf, 6v Capacitor	US Semcor
C51	X22	.022 uf, 15v Capacitor	Components
C52	L47	47 uf, 2v Capacitor	Components
C53	M6R8	6.8 uf, 10v Capacitor	Components
C54-55	L22	22 uf, 10v Capacitor	Components
C56-57	WM12BX103M	.01 uf, 100v Capacitor	Westcap
C58	TS02-6-505	5 uf, 6v Capacitor	US Semcor
C59	X-22	.022 uf, 15v Capacitor	Components
C60	L47	47 uf, 2v Capacitor	Components
C61	M6R8	6.8 uf, 10v Capacitor	Components
C62-63	L22	22 uf, 10v Capacitor	Components
C64-65	WM12BX103M	.01 uf, 100v Capacitor	Westcap
C66	TS02-6-505	5 uf, 6v Capacitor	US Semcor
C67	X-22	.022 uf, 15v Capacitor	Components
C68	L47	47 uf, 2v Capacitor	Components
C69	M6R8	6.8 uf, 10v	Components
C70	390-000X5U0471M	470 pf Capacitor	Erie
C71	CD6-100J	10 pf Capacitor	Cornell Dubilier
C72	WM12BX103M	.01 uf Capacitor	Westcap
C73	2954	0.8-10pf Var. Capacitor	Johanson
C74	TS01-10-105	1 uf, 10v Capacitor	US Semcor
C75	390-000X5U0471M	470 pf, Capacitor	Erie

C76	CD6-101J	100 pf, Capacitor	Cornell Dubilier
C77-78	L22	22 uf, 10v, Capacitor	Components
CR1-8	T1-6	Diode	Texas Instruments
CR9	GMD-1	Diode	Transitron
CR10-16	T1-6	Diode	Texas Instruments
CR17	GMD-1	Diode	Transitron
CR18-22	T1-6	Diode	Texas Instruments
CR23	GMD-1	Diode	Transitron
CR24-29	T1-6	Diode	Texas Instruments
CR30	GMD-1	Diode	Transitron
CR31-36	T1-6	Diode	Texas Instruments
CR37	GMD-1	Diode	Transitron
CR38-41	T1-6	Diode	Texas Instruments
CR42-43	PD6000	Zener Diode	PSI
CR44-45	GMD-1	Diode	Transitron
CR48-53	GMD-1	Diode	Transitron
CR54	1N4099	Zener Diode	Motorola
Q1-6	2N2206	Transistor	RCA
Q7	2N2419	Unijunction Transistor	GE
Q8-9	2N3391A	Transistor	GE
Q10	MM999	Transistor	Motorola
Q11-12	2N2206	Transistor	RCA
Q13	A306	Transistor	Amperex
Q14-15	2N3391A	Transistor	GE
Q16	MM999	Transistor	Motorola
Q17	2N3391A	Transistor	GE
Q18	MM999	Transistor	Motorola
Q19-21	2N3391A	Transistor	GE
Q22	MM999	Transistor	Motorola
Q23-26	2N3391A	Transistor	GE
Q27	MM999	Transistor	Motorola
Q28-31	2N3391A	Transistor	GE
Q32	MM999	Transistor	Motorola
Q33-36	2N3391A	Transistor	GE
Q37	MM999	Transistor	Motorola
Q38	2N3294	Transistor	Motorola
R1	MF4C-100K	100K, 1% Resistor	Electra
R2		1.5K, 10%, 1/10W Resistor	Ohmite
R3-13		15K, " " "	Ohmite
R4-14		100K, " " "	Ohmite
R15		1K, " " "	Ohmite
R16		100K, " " "	Ohmite
R17-18		22K, " " "	Ohmite
R19		27K, " " "	Ohmite
R20		150K, " " "	Ohmite
R21		390K, " " "	Ohmite
R22		150K, " " "	Ohmite
R23-36		180K, " " "	Ohmite
R37		18K, " " "	Ohmite
R38-39		15K, " " "	Ohmite
R40		6.8K, " " "	Ohmite
R41		47K, " " "	Ohmite
R42		100 ohms,10%, 1/10W Resistor	Ohmite
R43		1K, " " "	Ohmite
R44-45		10K, " " "	Ohmite
R46		47K, " " "	Ohmite

R47		22K, 10%, 1/10W Resistor	Ohmite
R48		4.7K, " " "	Ohmite
R49		1K, " " "	Ohmite
R50		15K, " " "	Ohmite
R51		680ohms, 10%, 1/10W Resistor	Ohmite
R53		470ohms, " " "	Ohmite
R54-55		220K, " " "	Ohmite
R56		100K, " " "	Ohmite
R57		33K, " " "	Ohmite
R58		220K, " " "	Ohmite
R59-60		47K " " "	Ohmite
R61		4.7K, " " "	Ohmite
R62-63		100K, " " "	Ohmite
R64		39K, " " "	Ohmite
R65-66		100K, " " "	Ohmite
R67		1K, " " "	Ohmite
R68		220K, " " "	Ohmite
R69		27K, " " "	Ohmite
R70	210-60-2.5K	2.5K Potentiometer	Daystrom
R71	210-60-50K	50K Potentiometer	Daystrom
R72-73		220K, 10%, 1/10W Resistor	Ohmite
R74		47K, " " "	Ohmite
R75		470ohms" " "	Ohmite
R76		100K, " " "	Ohmite
R77		33K, " " "	Ohmite
R78		220K, " " "	Ohmite
R79	MF4C-100K	100K, 1%, Resistor	Electra
R80		100K, 10%, 1/10W Resistor	Ohmite
R81	MF4C-100K	100K, 1%, Resistor	Electra
R82		68K, 10%, 1/10W Resistor	Ohmite
R83		10K, " " "	Ohmite
R84		8.2K, " " "	Ohmite
R85		100K, " " "	Ohmite
R86		220K, " " "	Ohmite
R87		100K, " " "	Ohmite
R88		220K, " " "	Ohmite
R89	210-60-1K	1K Potentiometer	Daystrom
R90		33K, 10%, 1/10W Resistor	Ohmite
R91		220K, " " "	Ohmite
R92		47K, " " "	Ohmite
R93		220K, " " "	Ohmite
R94		100K, " " "	Ohmite
R95		1K, " " "	Ohmite
R96	MF4C-100K	100K, 1% Resistor	Electra
R97		100K, 10%, 1/10W Resistor	Ohmite
R98	MF4C-100K	100K, 1% Resistor	Electra
R99		8.2K, 10%, 1/10W Resistor	Ohmite
R100		100K, " " "	Ohmite
R101		220K, " " "	Ohmite
R102		100K, " " "	Ohmite
R103		220K, " " "	Ohmite
R104	210-60-1K	1K Potentiometer	Daystrom
R105		33K, 10%, 1/10W Resistor	Ohmite
R106		220K, 10%, 1/10W Resistor	Ohmite
R107		47K, " " "	Ohmite
R108		220K, " " "	Ohmite

R109	MF4C-100K	100K, 1% Resistor	Electra
R110		100K, 10%, 1/10W Resistor	Ohmite
R111	MF4C-100K	100K, 1% Resistor	Electra
R112		8.2K, 10%, 1/10W Resistor	Ohmite
R113		100K, " " "	Ohmite
R114		220K, " " "	Ohmite
R115		100K, " " "	Ohmite
R116		220K, " " "	Ohmite
R117	210-60-1K	1K Potentiometer	Daystrom
R118		33K, 10%, 1/10W Resistor	Ohmite
R119		220K, " " "	Ohmite
R120		47K, " " "	Ohmite
R121		33K, " " "	Ohmite
R122		10K, " " "	Ohmite
R123		1K, " " "	Ohmite
R124		100K, " " "	Ohmite
L1		Coil Transmitter	Repco
T1		Transformer	Repco
T2		Transformer	Repco
J1	SMRE14S-J	Connector, 14 Pin	Winchester
P1	SMRE14P-JT	Connector, 14 Pin	Winchester

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The research and development described in this report resulted in the redesign and miniaturization of a Personal Telemetry Transmitter System originally developed by AMRL. The Personal Telemetry Transmitter System transmits seven channels of physiological data on the commercial FM band to a receiver located up to 200 feet away. The seven channels transmitted by pulse duration modulation are electroencephalogram, two leads of electrocardiogram, galvanic skin resistance (base resistance and specific response), respiration, and body temperature. Extremely compact packaging combined with miniature components resulted in a package size of 4.28 by 2.19 by .81 inches for a total volume of 7.6 cubic inches including the battery.

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Physiology, telemeter systems						
Space medicine						
Electronic equipment, miniaturized (man portable)						
Redesign						
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