A MULTICHANNEL PHYSIOLOGICAL DATA/VOICE TELEMETRY SYSTEM

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William M. Portnoy, Principal Investigator

Alva E. Henderson, Research Assistant

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Biomedical Instrumentation Laboratory

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CHAPTER I

INTRODUCTION

Telemetry consists of performing measurements at a remote location and reproducing these measurements at some convenient location in a form suitable for display and recording. The link connecting the two locations may consist, for example, of a sound wave, a modulated light beam, a telephone line, or a radio.[1] In this context, the first physiological telemetry system was developed by the U.S. Army Signal Corps in 1921. The system was to be used for transmission of heart sounds from ships without onboard physicians to medical facilities on shore. The system consisted of a microphone which rested on the patient's chest, and an audio amplifier.[2] Very little additional work was done in physiological telemetry until 1949, when an FM radio link was used in the telemetry of human electroencephalograms (EEGs). Breaksell and Parker had realized the possibilities of observing electrophysical data remotely and constructed the first modern working system.[2]

Since 1949, a great number of physiological radio telemetry systems have been developed. These range from large complex multi-channel, multi-patient monotoring systems for hospital use[3] to battery powered multi-channel portable units for such uses as exercise physiology[4] to small single channel implantable systems for biological experimentation.[5] With advances made in electronic integrated circuits, systems requiring less power and space, yet having greater capability, are emerging.

B. Purpose and Scope of Work

The physiological telemetry system described in this report was designed for a space environment under contract with the National Aeronautics and Space Administration Lyndon B. Johnson Space Center (NASA JSC). The desired result of the work was to deliver a prototype radio telemetry system which would transfer all required information (physiological as well as voice) to and from the astronaut. The actual transmission of information includes interfacing with the space vehicle's microwave link. This effectively reduces the required RF range of this prototype telemetry system.

For use in a remote environment, such as in outer space, reliability and flexibility of operation take on increased importance. The system's reliability must be such that no adjustments are required of the astronaut or of personnel on ground. The flexibility of the system must be such that either the astronaut or ground control has the capability to receive or transmit information as it is needed. Additional environmental constraints include a small self-contained portable unit which has minimum power requirements.

In recent NASA space missions, there have been occasions where, for some reason or other, the astronaut has turned his physiological and voice systems off. Obviously, this caused some anxious moments at ground control. As a result, it was emphasized that the prototype system must circumvent this problem by transferring control of the portable system from the astronaut to ground control. In addition, this transfer of control from the astronaut leaves him free to use his hands for other operations.

In any telemetry system, the data are subject to error causing noise in the transmission channel and non-linear effects in transmitters and receivers which distort the actual measured data. The degree of error introduced by these effects depends on the type of modulation used. The improved signal-to-noise ratio of Frequency Modulation (FM) over other modulation schemes is well known.[6] However, the use of FM does not circumvent the problem of nonlinear effects in FM transmitters and receivers. For telemetry of clinical quality physiological measurements, increased accuracy of the telemetry system is required.

The use of compound modulation systems such as FM-FM and PCM-FM modulation is particularly effective. By converting the measured data into a digital signal (PCM), errors introduced by nonlinear effects are reduced enormously. Using the PCM signal as the modulating signal in a FM transmitter, the compoundly modulated signal (PCM-FM), is immune to both nonlinear and noise induced errors. For the above reasons, PCM-FM is well suited for telemetry systems requiring high accuracy and resolution of the measured data.

Advances made in solid state integrated components make a low power, complex, telemetry system possible. CMOS has ultra-low power requirements. For medium speed operation, typical power comsumption of a medium scale CMOS digital circuit is less than 10 μwatts.[7] Low power operational amplifiers are also available which operate with quiescent currents as low as 50 μa.[8] This, coupled with the short range required of the RF links, makes a low power system with good reliability practical.

A complete description of the designed physiological data telemetry system is given in Chapter II. In addition to describing the various modes of operation and interaction of the different subsystems, design philosophy is also discussed. This discussion includes difficulties encountered and the engineering compromises that resulted. In Chapter III, specific electronic circuit design of the subsystems is given. Chapter IV provides details of the actual system performance. Also included in Chapter IV are suggested modifications that will more easily lend the system to hybrid construction and that will circumvent some basic problems discussed in Chapters II and III.

radio frequency interference, respectively. Other parameters were left unspecified to allow some flexibility in implementation.

B. General System Description

Basically, the system consists of a ground based station (GBS) and a portable unit worn by the astronaut (Figure 1). These two units are connected by two separate RF links. Two separate RF links are required to implement control and to facilitate low power consumption. The portable unit is composed of a receiver subsystem that receives voice and control signals from GBS; a PCM encoder subsystem that digitizes physiological and calibration data from the astronaut; a transmitter subsystem that transmits voice and PCM data to GBS; and an interfacing control subsystem. The GBS unit is composed of a receiver that receives voice and PCM data from the astronaut; a PCM decoder that reconstructs the physiological data as it is received; a transmitter that transmits voice and control information to the portable unit; and a control signal generator. A block diagram of the system is given in Figure 1.

C. Control System

As emphasized before, control of the system must originate at GBS. A control subsystem is required which will permit two distinct PCM modes of operation:

1. an operate mode, consisting of instruction to the portable unit to send physiological data;
2. a calibrate mode, consisting of instruction to the portable

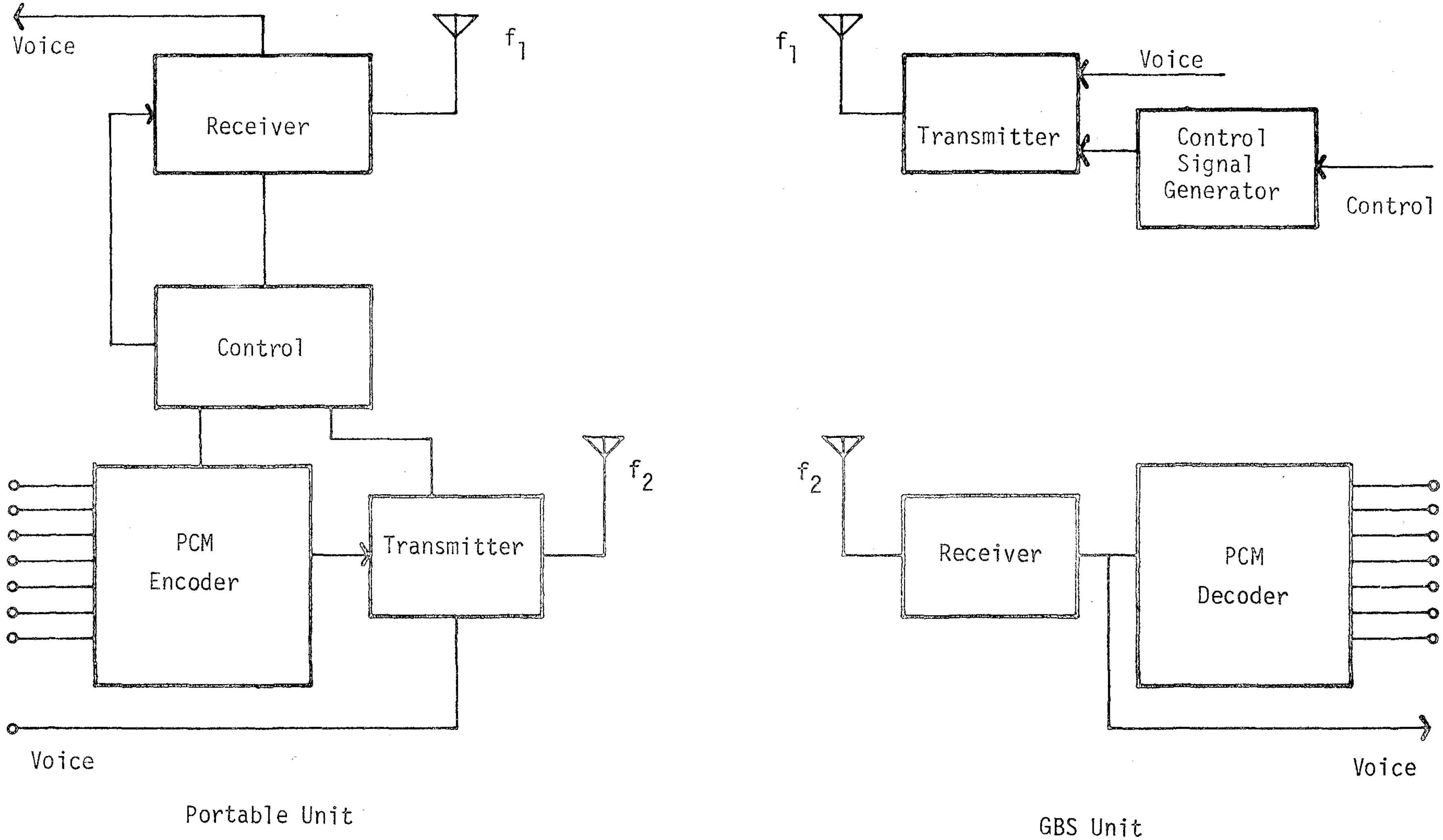


Figure 1. System Block Diagram.

unite to provide a calibrated square wave to determine the gains of the individual physiological channels.

In addition, remote control reduces power consumption by switching power only to the required subsystem. To implement this type of control, the portable unit requires a control signal from GBS regardless of the state of the portable unit transmitter. Consequently, two separate RF channels with capability for simultaneous transmission are required.

In the early design stages it was intended that transmission of the astronaut's voice and PCM data be made simultaneously. However, as will be discussed later, this was not possible using a single RF link to GBS, within NASA specifications. Consequently some control by the astronaut was included. Astronaut control consists of a voice activated switch (VOX) which turns on power to the transmitter automatically as he speaks. Additionally, if PCM data are being transmitted, the astronaut has temporary interrupt control. By activating the interrupt control, the astronaut disables PCM transmission so that he may establish voice contact to GBS. This arrangement prevents permanent termination of any part of the portable unit.

The control signal transmitted by GBS consists of a two bit digital tone encoded control word. The two bit format allows four possible control instructions. Two of these instructions are reserved for operate and calibrate commands. The third instruction is used for the condition in which no PCM data is required, but in which the RF links are needed for voice communications. The last instruction signifies that no control signal is being transmitted and that no part of the portable unit requires activation (contingent on the astronaut's control).

The total control format consists of eight control instructions which address six distinct states of subsystem power control (Figure 2). Redundancy of subsystem power states is the result of dual PCM interrupt instructions by both astronaut and GBS controller.

B. Portable Unit

1. Receiver

The receiver located in the portable unit operates in a sampling mode when no control signal is present. That is, power is switched to the receiver during relatively short intervals (300 msec), so that it may detect a change in GBS control instruction. If a change is detected, the receiver is locked on and remains on until the control signal is terminated by GBS. If no GBS control signal is transmitted during the sampling period, the receiver turns off and remains off for a relatively long period (2 sec). After the 2 second off period, the receiver is turned on again for the 300 msec sampling period. This type of cyclic receiver operation requires low average power consumption and consequently extends battery life.

The receiver is basically a narrowband crystal controlled FM receiver (Figure 3). Double superheterodyning converts the carrier frequency of 154.570 MHz to a 455 KHz IF signal. A phase-locked loop detector follows the IF stage. The detected signal is fed to two tone decoders which detect GBS control instructions and to an audio

Tones		VOX	INT	TRANS	REC	Encoder	
T_1	T_2					CAL	OPER
0	0	0	0	0	0	0	0
X	X	0	1	0	1	0	0
0	0	1	0	1	0	0	0
X	X	1	1	1	1	0	0
0	1	X	0	1	1	0	1
1	0	X	0	1	1	1	0
1	1	0	0	0	1	0	0
1	1	1	0	1	1	0	0

1 - indicates signal

0 - indicates no signal

X - indicates a don't care condition

1 - indicates power switched to respective subsystem

0 - indicates no power

Figure 2. Control Truth Table

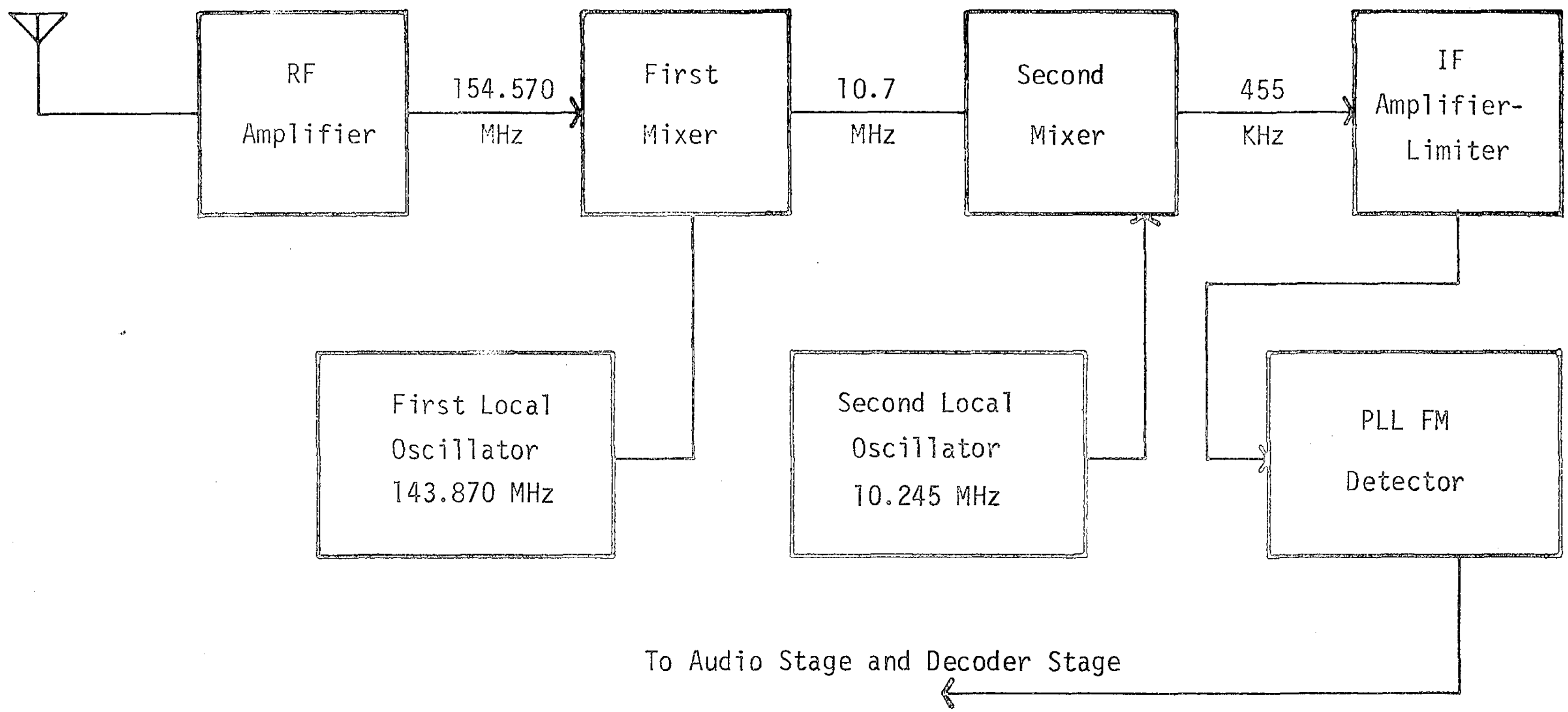


Figure 3. Portable Unit Receiver Block Diagram.

stage which amplifies voice.

Originally a single conversion receiver with a 10.7 MHz IF frequency was designed. However, inadequate selectivity and sensitivity of the PLL detector operating at 10.7 MHz, established the requirement for a lower IF frequency. At a center operating frequency of 455 KHz, narrowband selectivity, good sensitivity, and very linear and stable operation of the PLL detector was obtained.

2. PCM Encoder

The PCM encoder subsystem converts seven channels of parallel physiological and calibration input data into a single serial output digital signal. Because of their small amplitude (ca. 1 mv), physiological signals require amplification prior to digital conversion. Consequently, instrumentation amplifiers (gain = 1000) are required for each physiological channel. Preceding each instrumentation amplifier is a switch network that selects physiological signal inputs or calibration signal inputs depending on GBS control instructions. The calibration signal generator generates a precision 1 mv square wave. The seven channels of amplified signals are then time multiplexed and converted to digital data (Figure 4).

The resulting digital data consist of time multiplexed data in NRZ (non-return to zero) format and appropriate synchronization data. The synchronization data include word synch and frame synch pulses (Figure 5). There are eight words per frame and 590 frames per second. This corresponds to 590 samples per channel per second, which is more than adequate for most physiological signals.[9] The eight words in each frame correspond to one channel of zero reference information, five channels of multiplexed data representing five physiological channels, one channel of sub-multiplexed data representing two channels of low frequency physiological data, and one word period for frame synchronization purposes. Each word is a 10 bit digital representation of the sampled value for a given channel. NRZ data and synchronization data are made distinguishable by the use of tri-state logic; that is, there are three levels of digital data represented by a +1, a 0, and a -1. A positive signal (+1) corresponds to NRZ data and a negative signal (-1) represents synchronization data (Figure 5).

The actual analog to digital conversion involves two steps. The first step requires conversion of the sampled amplitudes into a pulse width modulated signal. A pulse width corresponding to a certain sample is then used to gate a precision crystal controlled oscillator to a counter. The digital number that is counted represents the value of the sampled amplitude. The primary purpose of this type of ADC implementation is lower power consumption.

3. Transmitter

The transmitter subsystem of the portable unit is a crystal controlled FM radio link, operating at 80 MHz. This RF frequency was chosen for a number of reasons:

1. adequate separation from the operating frequency of the receiver in the portable unit;
2. frequency is sufficiently high so that small efficient

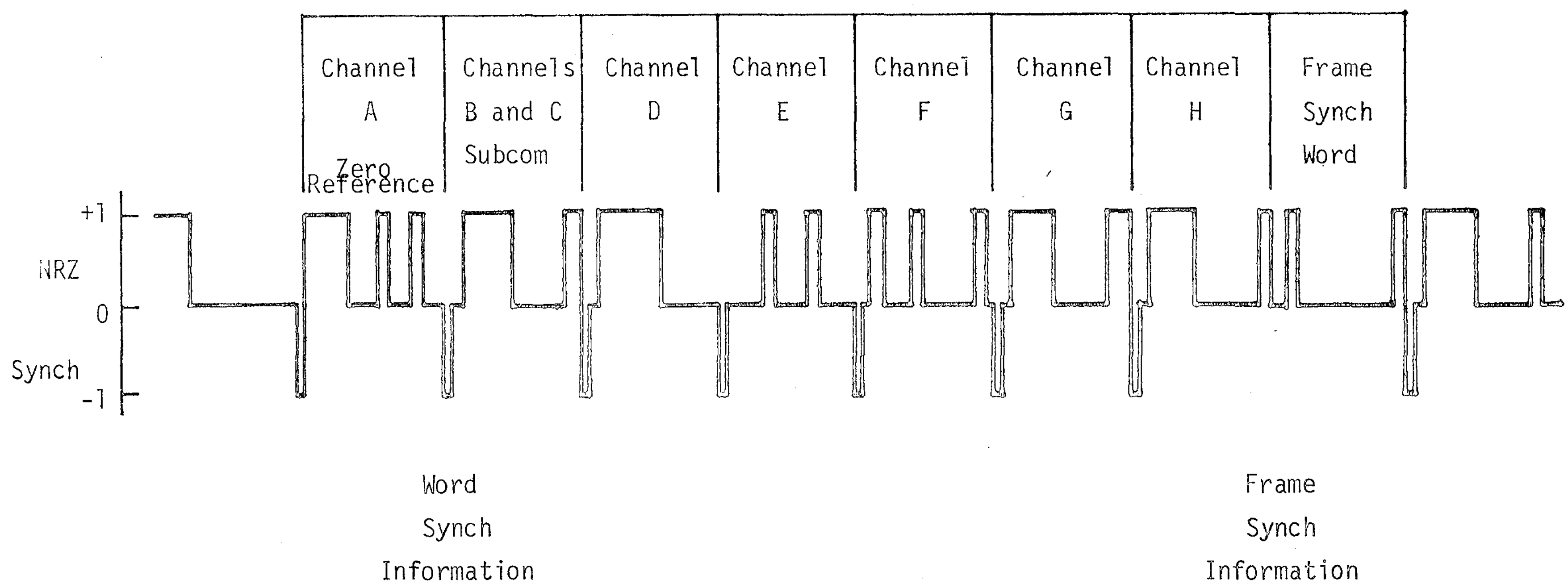


Figure 5. PCM Format.

antennae are realizable; and

3. local RF interference is minimal.

The transmitter must transmit both astronaut voice signals and PCM data signals. The two signals are considerably different in nature. Ordinary voice transmission requires a bandwidth between 300 Hz and 3 KHz. However, the PCM data has considerably greater bandwidth. As illustrated in Figure 5, a single bit of PCM data has a period of 19.2 sec which corresponds to a bit rate frequency of approximately 50 KHz. Because of the square waveshape of the PCM data, synchronization data, and the varying PCM pulse duration (NRZ format), the actual frequency spectrum of the PCM signal has a bandwidth from below 500 Hz to over 100 KHz. For good signal-to-noise ratio, a modulation index of at least one is required.[6] The wide bandwidth of the PCM signal, coupled with crystal control, causes implementation problems.

Two methods for implementing simultaneous transmission of two signals on one FM link are:

1. addition of the two signals linearly and use of the composite signal to modulate a FM transmitter;

2. sideband operation for one signal.

For the first method to be useful, the frequency spectrum of the two signals must not overlap. If they do overlap, proper separation by filtering would be impossible. Generation of an FM sideband (second method) requires a more sophisticated transmitter and consequently more power consumption and more space.

Crosstalk that exists between voice and PCM data rules out the first method. The increased complexity of implementation required by the second method is also prohibitive within the given specifications of low power and small size. As an engineering compromise, a temporary PCM data interrupt control was added.

A modulation index of at least one for both voice and PCM data requires approximately a 5 KHz deviation for voice and a 50 KHz deviation for PCM data. The modulator that was designed to provide these deviations consisted of a varactor diode controlled direct FM stage for voice, cascaded to an indirect (phase modulation) FM stage for PCM data. The use of both direct and indirect FM modulation was engendered by the inherent bandwidth (or deviation) differences between frequency and phase modulation. The deviation is independent of modulation frequency for FM, whereas the deviation is linearly related to the modulating signal frequency for PM.[10]

The linear relationship between deviation and modulating signal frequency in phase modulation, together with the difficulty of designing wide deviation crystal controlled direct FM modulators, makes phase modulation more suitable for producing wide frequency deviations. It must be emphasized, however, that phase modulation has the disadvantage of extremely wide bandwidth corresponding to the higher harmonics related to the square wave shaped PCM data. Converting phase modulation to indirect FM requires a $1/f_m$ (f_m denotes frequency of modulating signal) attenuating network preceeding the phase modulator to offset the linear increase of deviation with f_m. This conversion, however, reduces deviation below what is required. Consequently a "pseudo-indirect" FM system was used.

The instantaneous phase angle of a phase modulated signal modulated by a sinusoid is expressed as

$$\Theta(t) = \omega_c t + K_p V_m \sin\omega_m t, \tag{1}$$

where ω_c is the carrier frequency, ω_m is the modulating signal frequency, and K_p represents the conversion gain of the phase modulator in radians per volt.[10]

The instantaneous frequency of a phase modulated signal is obtained by differentiating Equation (1) with respect to time:

$$\omega = \frac{d\theta}{dt} = \omega_c + K_p V_m \omega_m \cos \omega_m t. \tag{2}$$

The peak deviation from the carrier frequency of ω_c is $K_p V_m \omega_m$.

If the modulating signal is passed through a low pass filter having an upper frequency cutoff of ω_1 Hz, the peak deviation will be

$$\omega_{peak} = K_p V_m \omega_1 \frac{\omega_m}{\omega_1 + s} \quad . \tag{3}$$

Replacing s with $j\omega_m$ (steady state), Equation (3) may be written as

$$\omega_p = K_p V_m \omega_1 \frac{\omega_m}{\omega_1 + j\omega_m} \quad . \tag{4}$$

Except for a 90 degree phase difference, the bracketed quantity in Equation 4 is identical to that which would be obtained using a high pass filter with a low frequency cutoff of ω_1 Hz. By choosing an appropriate value of ω_1, a compromise between deviation (which increases linearly with ω_1) and attenuation of low frequency information (which also increases with ω_1) is obtained. If low frequencies are not important, the value of ω_1 may be made larger and consequently more deviation may be obtained without excessive loss of information. To distinguish this scheme from a strict indirect FM scheme the term "pseudo-indirect" FM is used.

Direct FM modulation is used for voice signal because the required modulation index is more easily obtained. For low modulation signal frequencies, phase modulation becomes very ineffective because of the low cutoff valve required of the low pass filter.

In addition to the cascaded modulators operating at 5.555 MHz, the transmitter consists of two frequency tripler stages and a frequency doubler stage. These tripler stages provide a factor of 18 increase in deviation and also provide a final transmitter frequency of 80 MHz (Figure 6).

E. Ground Based Station (GBS)

1. Receiver

The receiver used in this work was a high sensitivity general purpose VHF receiver, with a frequency range between 30 MHz and 300 MHz (Airborne Instruments Laboratory Model R-1283/GRC). It also had selectable IF bandwidths of 60 KHz, 300 KHz, and 3 MHz. For prototype

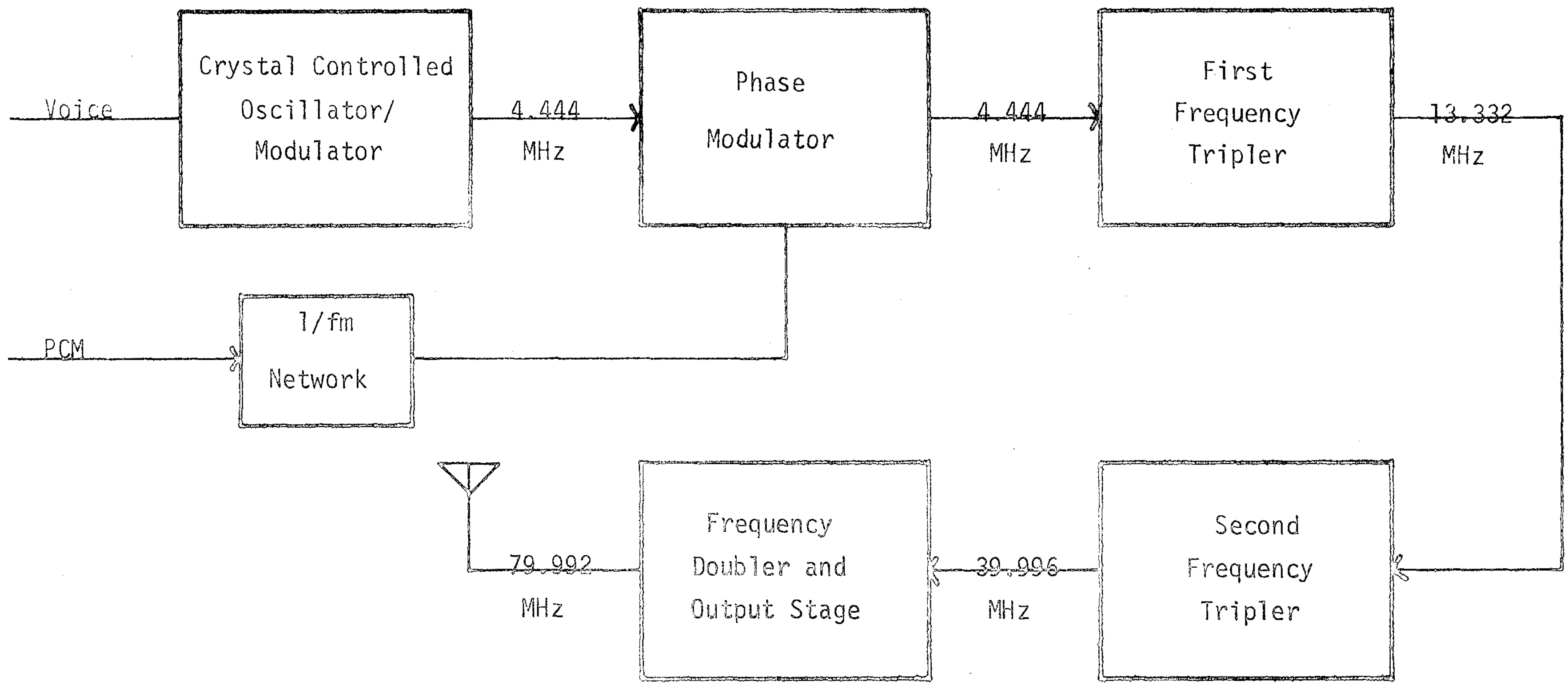


Figure 6. Portable Unit Transmitter Block Diagram

to encode the two bit control instruction word. The tone reeds stabilize the sine wave frequency to within 0.15%. The actual sine wave generator is a commercial unit (Repco Inc. Part Number 810-024-02) compatible with the GBS transmitter. A modification of the tone generator was required, however, to implement a full two bit operation, since the tone generator would not accept both reeds simultaneously. To simulate simultaneous tone transmission, the two tone reeds were alternately transmitted at a switching rate of approximately 1 Hz.

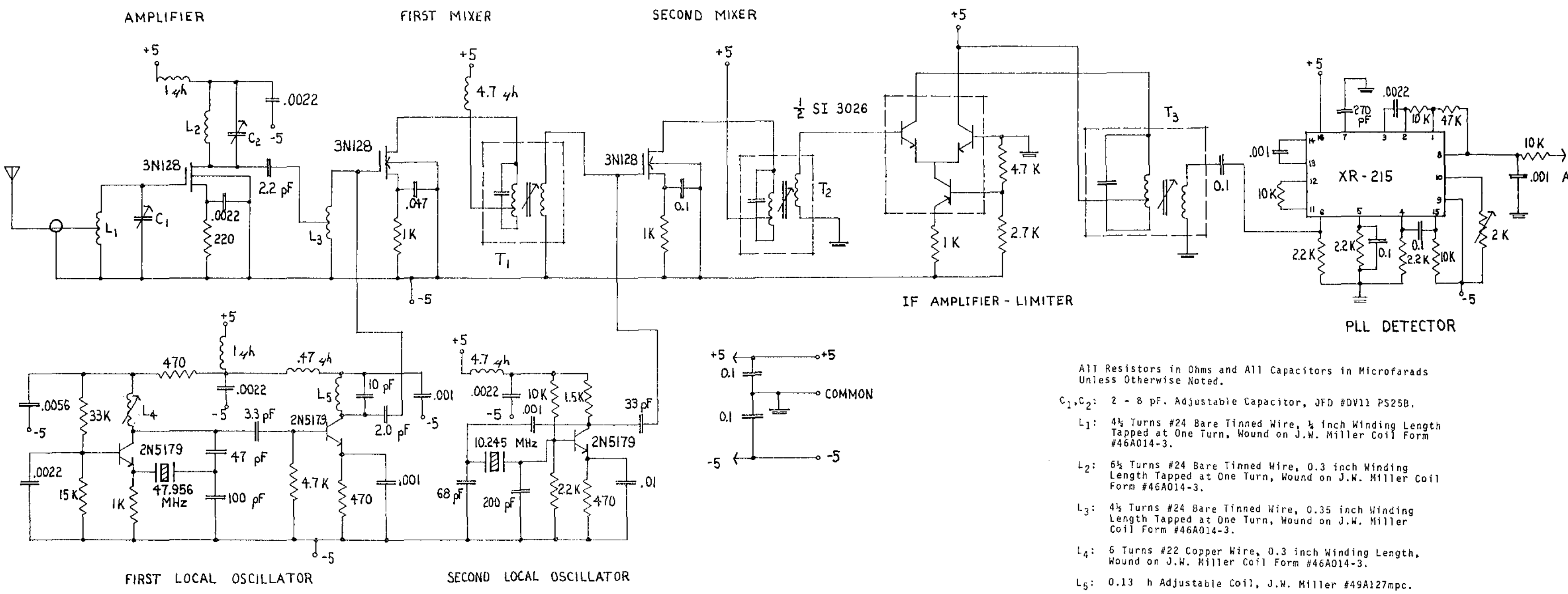
±3.75 volt power source. The operation of the phase-locked loop detector in the portable unit receiver, however, becomes marginal at this voltage. Consequently, two cells were added to provide a ±5 volt supply so that proper operation of the phase-locked loop detector would be ensured.

B. Portable Unit

1. Receiver

The receiver consists of an RF amplifier, two cascaded mixer stages, a single IF amplifier-limiter stage, and a phase-locked-loop detector (Figure 7). The RF amplifier and mixer stages employ MOS transistors as the active elements. MOS transistors exhibit almost ideal square law behavior; consequently, good mixing action and low generation of spurious harmonics result.[10] Receiver sensitivity is not a critical factor since the receiver must operate satisfactorily at only a distance of 200 to 300 feet from a 0.5 watt transmitter source. More than adequate sensitivity is obtained by a single IF amplifier stage.

The MOS transistor in the RF amplifier is biased at a quiescent current of approximately 2 ma. This value represents a compromise between power consumption and power gain. Additional power gain can be achieved by higher bias current; however, the rate of increase of power gain with respect to power dissipation becomes noticeably smaller beyond a bias current of 2 ma.[10] The RF input tank circuit is designed to provide conjugate match between the 50 ohm antenna source and the input impedance (Y_{11} = 0.45 + j5.57 millimohs) of the



All Resistors in Ohms and All Capacitors in Microfarads Unless Otherwise Noted.

C_1, C_2: 2 - 8 pF. Adjustable Capacitor, JFD #DV11 PS25B.

L_1: 4½ Turns #24 Bare Tinned Wire, ¼ inch Winding Length Tapped at One Turn, Wound on J.W. Miller Coil Form #46A014-3.

L_2: 6½ Turns #24 Bare Tinned Wire, 0.3 inch Winding Length Tapped at One Turn, Wound on J.W. Miller Coil Form #46A014-3.

L_3: 4½ Turns #24 Bare Tinned Wire, 0.35 inch Winding Length Tapped at One Turn, Wound on J.W. Miller Coil Form #46A014-3.

L_4: 6 Turns #22 Copper Wire, 0.3 inch Winding Length, Wound on J.W. Miller Coil Form #46A014-3.

L_5: 0.13 h Adjustable Coil, J.W. Miller #49A127mpc.

T_1: 10.7 MHz Transformer, J.W. Miller #8856.

T_2: 455 KHz Transformer, J.W. Miller #8815.

T_3: 455 KHz Transformer, J.W. Miller #8816.

Figure 7. Portable Unit Receiver Schematic

FOLDOUT FRAME 1

ORIGINAL PAGE IS OF POOR QUALITY

FOLDOUT FRAME 2

MOS transistor at 154.57 MHz. The output circuit is designed to match the output impedance (Y_{22} = 0.28 + j1.35 millimhos) of the MOS transistor at 154.57 MHz to a 50 ohm load. Neutralization was not used because additional gain was not required.

The first local oscillator operates at a frequency of 143.87 Mhz and is crystal controlled. The oscillator consists of a 47.956 MHz crystal oscillator cascaded to a frequency tripler. The 47.956 crystal oscillator is a common base Colpitt oscillator with the crystal operating in series resonance.[11]

The first mixer stage employs a MOS transistor at a bias current of approximately 1.5 ma. At this bias, the transistor transconductance exhibits the greatest amount of nonlinear (square law) behavior.[12] The input gate circuit is designed for a conjugate match of the 50 ohm RF signal source with the input impedance of the MOS transistor. The first local oscillator signal is injected to the gate through a small (3 pf) coupling capacitor. The output tank circuit is a tuned transformer resonant at 10.7 MHz which filters out the difference frequency.

The secondary of the 10.7 MHz transformer is connected directly to the gate of the second mixer stage. The transformer has a turns ratio of seven to one. For a match between the output of the first mixer and the input to the second mixer a turns ratio of two to one is required. A mismatch was designed deliberately in order to increase the Q factor of the transformer and consequently improve selectively. As in the first mixer, a bias current of 1.5 ma was employed. The output circuit of the second mixer is a tuned transformer resonant at the difference frequency of 455 KHz and having a turns ratio of five to one.

The second local oscillator is a Pierce crystal controlled oscillator operating at 10.245 MHz. The crystal operates in the parallel resonance mode. The signal generated by this oscillator is injected to the gate of the second mixer by means of a small (15 pf) coupling capacitor.

The secondary of the 455 KHz transformer is connected to a single IF amplifier-limiter stage. Again, a mismatch is used to increase selectivity. The IF amplifier-limiter stage consists of a differential amplifier with constant current bias. This configuration permits non-saturating limiting operation. The IF amplifier is operated from a split power supply for proper differential operation. A constant current bias of approximately 1 ma provides a full limiting output of approximately 0.5 volt RMS across the phase-locked loop detector input.

The phase-locked loop demodulator circuit is essentially the circuit described for split power supply operation in the EXAR 1972 EX-215 phase-locked loop data sheet.[13] Modifications of this circuit include a change in center operating frequency and a change in lock range (selectivity). The PLL detector was designed for a center frequency of 455 KHz and a bandwidth or lock range of 25 KHz. For good stability of the VCO center frequency, a stable silver mica capacitor was used for the VCO timing capacitor. The range extension resistor is made variable to permit frequency adjustment. The demodulated signal at the output of the preamplifier located on the XR-215 integrated circuit is fed to a buffer stage and subsequently to the tone decoders and an audio stage.

The detection threshold input voltage for the XR-215 PLL is approximately 3 mv RMS; however, for consistent lock range characteristics, the input voltage should be at least 30 mv RMS. A conservative calculation of the power gain of the RF front end and IF amplifier stages is 60 db. The receiver sensitivity for consistent receiver operation is therefore at least 30 microvolts. This sensitivity is more than adequate for the application.

2. Control and Audio Circuits

The audio circuits are included with the control circuits in this section because of the interrelation of the two. The interrelation between voice and control exists on both voice channels. On the astronaut voice channel, the control signal VOX depends directly on the astronaut's audio signal. GBS control information and GBS voice information are related in that they are transmitted over the same RF link.

One of the two tone decoders which follow the receiver consists of a bandpass filter followed by a peak detector and voltage comparator (Figure 8). A design outline for this circuit is presented in the Siliconix application note for the L144 operational amplifier. The active bandpass filter is a dual integrator feedback resonator. The filter implementation is preferred over multiple feedback because

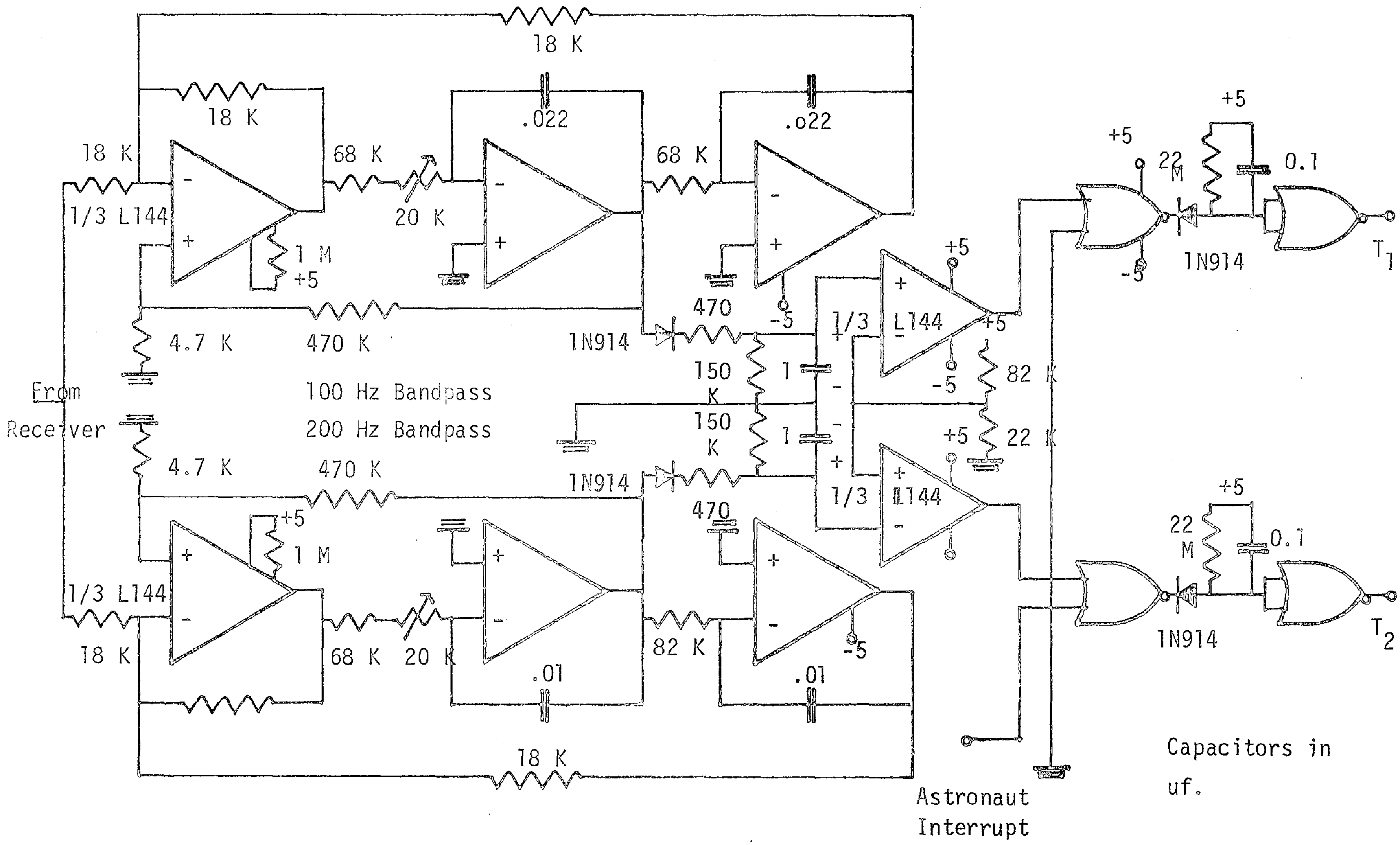


Figure 8. Tone Decoder Schematic.

of the low sensitivity to changes in component values and ease of obtaining center frequency adjustment.[8] The RC time constants of the two integrators are related to the bandpass characteristics in the following ways:

$$R_2C_2 = \frac{H_o}{2\pi f_o Q} \quad \text{and} \tag{5}$$

$$R_1C_1 = \frac{Q}{2\pi f_o H_o}, \tag{6}$$

where Q describes the bandwidth of the filter, H_o denotes the center frequency gain of the filter, and f_o denotes the filter center frequency. The subscripts 1 and 2 correspond to the first and second integrator, respectively.

By assuming that $H_o = Q = 30$, Equations 5 and 6 become equivalent, so that, if $R_1 = R_2 = R$ and $C_1 = C_2 = C$, then

$$RC = \frac{1}{2\pi f_o}, \tag{7}$$

If R_1 is less than R and R_2 is greater than R, the assumption that $Q = H_o$ is false. However, if R_1 is made variable, the relationship between the two RC time constants and the H_o/Q and Q/H_o ratios establishes a procedure for center frequency adjustment.

The two bandpass filters were designed for center frequencies of 100 Hertz and 200 Hertz, corresponding to the tone frequencies. A Q of 25 and an H_o of 30 were designed for adequate selectivity and sensitivity for following stages. An integrated tone decoder requiring only one external resistor and one external capacitor is commercially available; however this circuit was not used because its high power consumption (11 ma).[14]

The tone decoders are followed by retriggerable one-shot multivibrators. These are required so that when the tones are alternately switched (GBS control instruction word 11), the decoded output is a constant level. The astronaut interrupt control is activated by causing both one shot multivibrators to lock in a high state. The state of the portable unit is thus controlled by the instruction signals T_1, T_2, and VOX, where T_1 and T_2 are the digital signals at the output of the retriggerable one-shot multivibrators.

A modified version of the truth table in Figure 2 (Chapter II) is given in Figure 9. The modification consists of combining the astronaut interrupt instruction with the GBS interrupt instruction (GBS control instruction word 11). The various commands and power switching instructions are related logically to the digital instructions T_1, T_2, and VOX by

$$P_{REC} = T_1 + T_2 \tag{8}$$

$$P_{ENC} = \overline{T}_1 T_2 + T_1 \overline{T}_2 \tag{9}$$

$$P_{TRANS} = (\overline{T_1 T_2} + T_1 T_2)VOX + \overline{T}_1 T_2 + T_1 \overline{T}_2 \tag{10}$$

$$\mathrm{OPERATE} = \overline{T}_1 T_2 \tag{11}$$

$$\mathrm{CALIBRATE} = T_1 \overline{T}_2 \tag{12}$$

Equations 8 through 12 are implemented as shown in Figure 10, which shows the complete schematic for the tone control logic and power switching circuits. For certain power switching instructions, an inverter was

INPUTS			OUTPUTS			
T_2	T_1	VOX	P_{REC}	P_{TRANS}	PCM	
					OPER	CAL
0	0	0	0	0	0	0
0	0	1	0	1	0	0
0	1	X	1	1	1	0
1	0	X	1	1	0	1
1	1	0	1	0	0	0
1	1	1	1	1	0	0

1 = TRUE

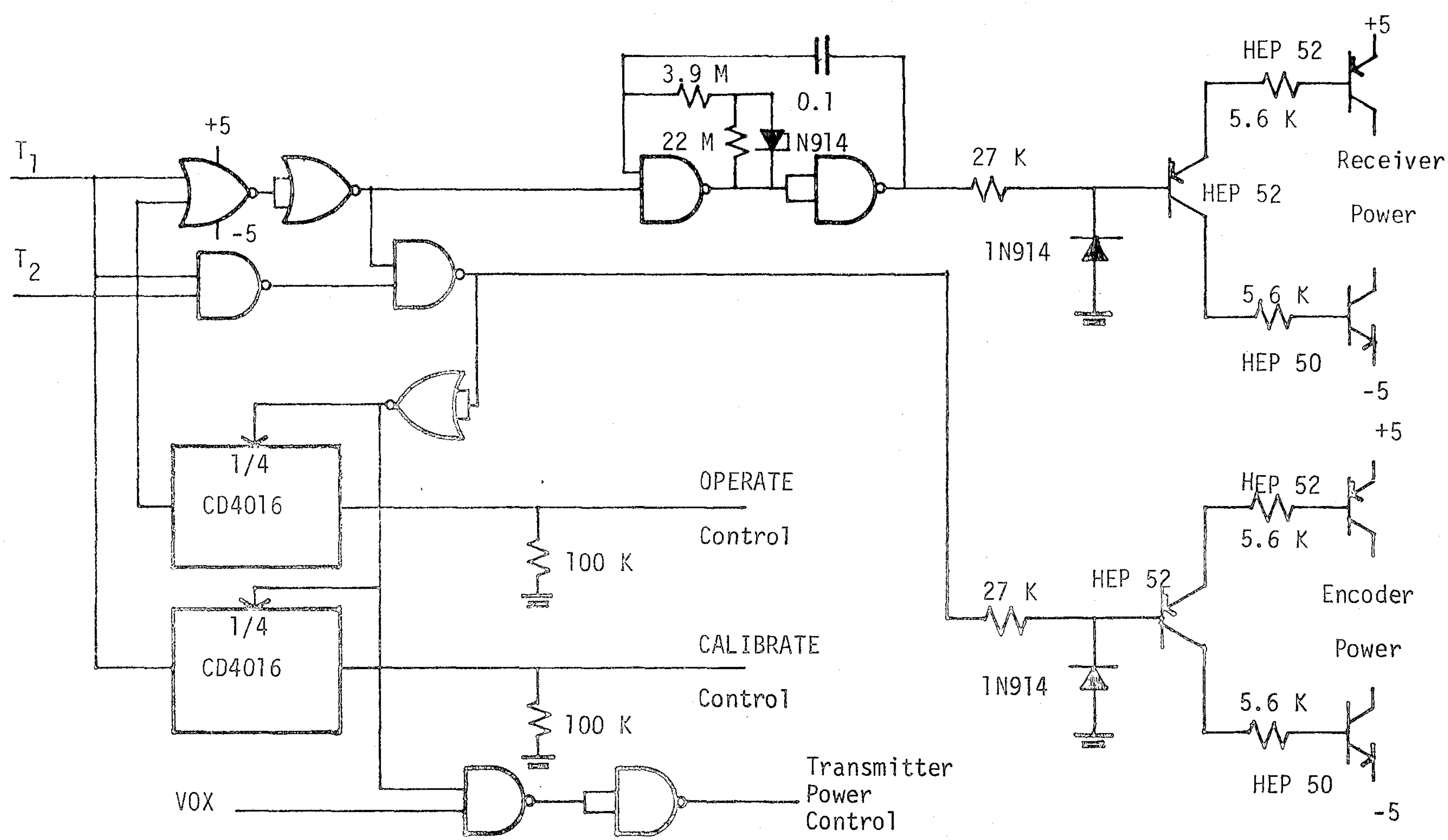
0 = FALSE

X = DON'T CARE

1 = POWER SWITCHED

0 = NO POWER SWITCHED

Figure 9. Tone Control Truth Table



Capacitors in uf.

Figure 10. Tone Control Logic Schematic.

required since a logical 0 turns the transistor switch on.

A free running (astable) multivibrator with short duty cycle operation (300 milliseconds ON and 2 seconds OFF) switches the receiver transistor power switch on and off in the sampling mode.[15] An interrupt input to this astable multivibrator is provided. A logical zero at this input allows free running operation, whereas a logical 1 (indication of tone signal transmitted) locks the receiver power on.

The transistor power switch uses both NPN and PNP transistors. This type of switch is required because of split supply operation. A logical 1 inhibits current flow to the respective subsystem. In the OFF state, the voltage applied to the subsystem is at ground or system common. A logical zero turns the transistors on and permits power flow to the subsystem load. For proper operation of the transistor switch, a diode is used at the input to clamp the logical 0 signal to approximately system common.

For safe operation of CMOS devices, the input signal must not exceed the power supply voltage.[16] However, if the control logic is such that power is not switched to the PCM subsystem, excessive input signal excursion due to the logical operate and calibrate commands can occur. To circumvent this problem, a CMOS switch was used to return OPERATE and CALIBRATE signals to system common when no power is delivered to the PCM subsystem.

The VOX control signal is generated using a peak detector and voltage comparator following the astronaut's voice amplifier (Figure 11). The resulting signal operates a retriggerable one-shot

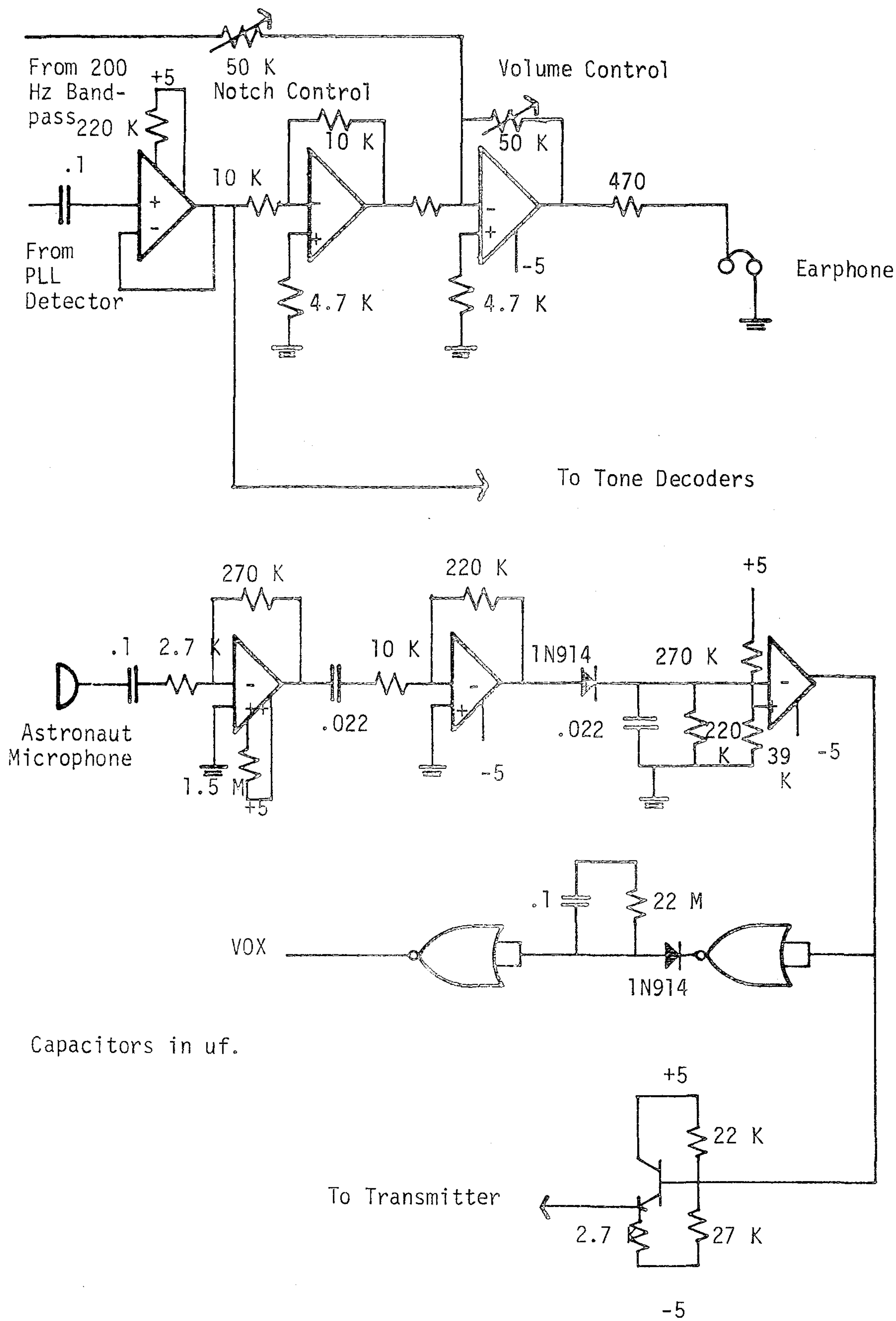


Figure 11. Portable Unit Audio Schematic.

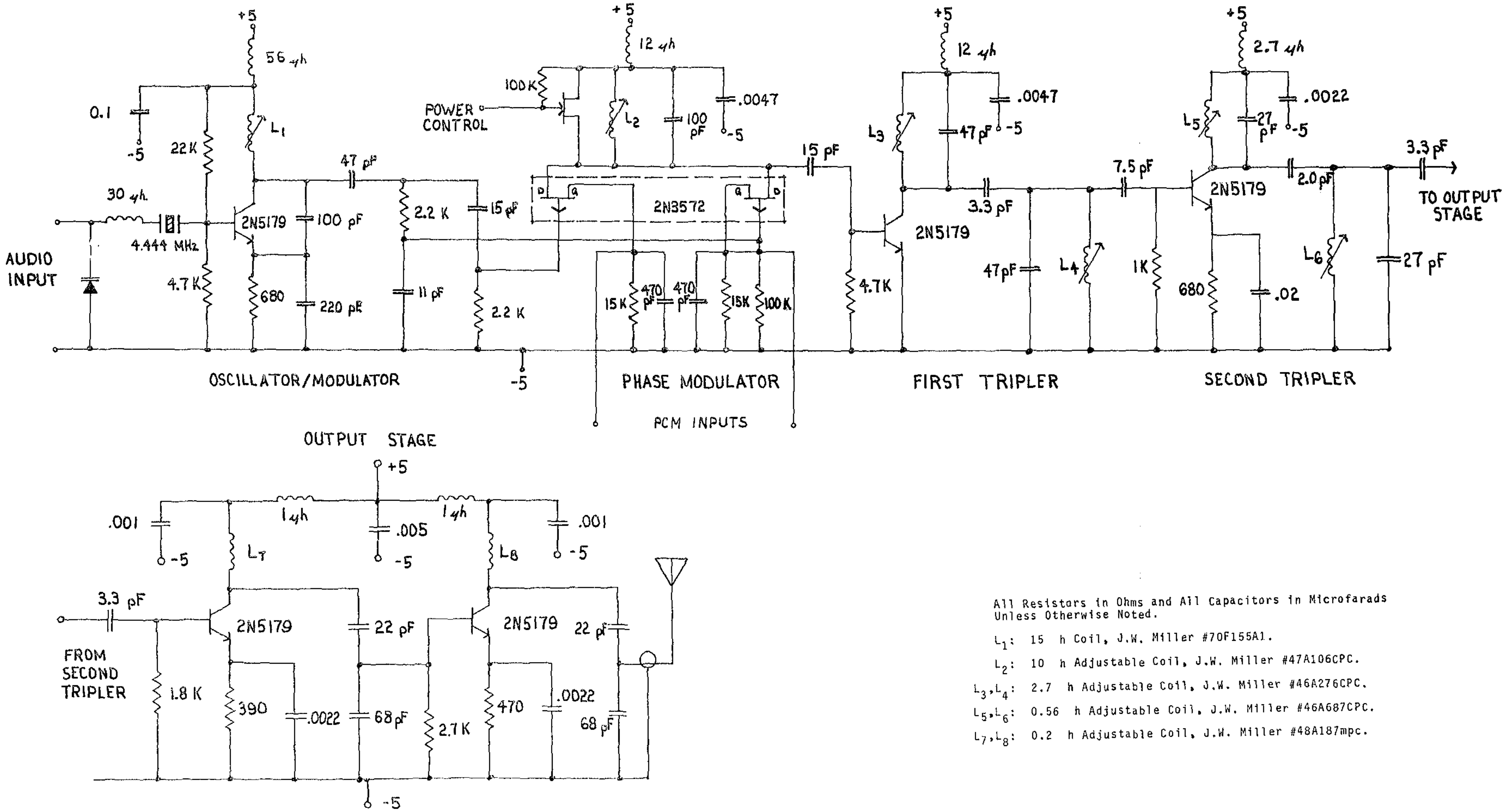
multivibrator with a one-shot time constant of approximately 2 seconds. This delay allows short pauses in voice signals while avoiding transients caused by power switching. The astronaut's voice amplifier consists of two cascaded operational amplifier stages. A gain of 2000 is required to amplify the small 1 millivolt microphone signal to one of sufficiently high amplitude to modulate the transmitter and to activate the VOX circuit. (A bone conduction microphone furnished by NASA JSC was used in this work.) Both stages are coupled to avoid DC offset at this high gain.

The 200 Hertz tone transmitted by GBS was originally assumed to be inaudible, but could actually be heard. To eliminate the tone, it was notched out in the receiver audio stages. The use of a notch filter requires a bandpass filter. To minimize the number of components, the output of the 200 Hertz bandpass filter in the 200 Hertz tone decoder was fed back to the audio stages (Figure 11). The second operational amplifier in the audio stage is used as a phase inverter to provide the proper phase relations for notch operation. The third operational amplifier has variable gain and drives the headset worn by the astronaut. The particular headset, provided by the NASA Johnson Space Center, was a low impedance type; consequently a current limiting resistor was used in series with it. This limiting resistor represents considerable signal loss; however enough power was delivered to the headset so that adequate volume resulted.

3. Transmitter

The first stage of the transmitter (Figure 12) consists of a crystal oscillator which doubles as a direct FM modulator. Transistor Q_1 operates in a common base configuration. The tank circuit is resonant at approximately the crystal frequency (4.444 MHz). Around 30 percent of the output signal is fed back to the emitter to sustain oscillations. The crystal, together with the inductance and capacitance in series with it, acts as a series resonant bypass element. Direct modulation is achieved by varying the bias across the varactor diode. A change of reverse bias across the varactor diode causes the equivalent capacitance to change, which subsequently causes the series resonance of the bypass branch to change. Small, yet adequate, deviation is possible with this circuit. The inductance in series with the diode and the crystal tends to linearize the capacitance-voltage characteristics of the varactor diode.[18]

The output signal of the oscillator-modulator is fed to a phase modulator. Phase modulation is obtained by adding two RF signals of fixed phase difference and of varying amplitude. The signal from the oscillator is applied to a phase splitting network. Two signals are produced: one leads the oscillator signal by 45 degrees and the other lags the oscillator phase by 45 degrees. In a push-pull type of operation, these signals are amplitude modulated by the modulating signal.[17] When these two signals are added together, the resulting signal is phase modulated. A graphical description of this procedure is shown in Figure 13. This push-pull type of phase modulation is



All Resistors in Ohms and All Capacitors in Microfarads Unless Otherwise Noted.

L_1: 15 h Coil, J.W. Miller #70F155A1.

L_2: 10 h Adjustable Coil, J.W. Miller #47A106CPC.

L_3,L_4: 2.7 h Adjustable Coil, J.W. Miller #46A276CPC.

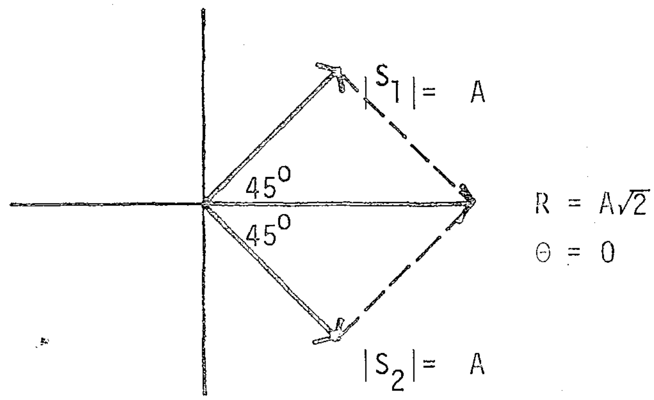
L_5,L_6: 0.56 h Adjustable Coil, J.W. Miller #46A687CPC.

L_7,L_8: 0.2 h Adjustable Coil, J.W. Miller #48A187mpc.

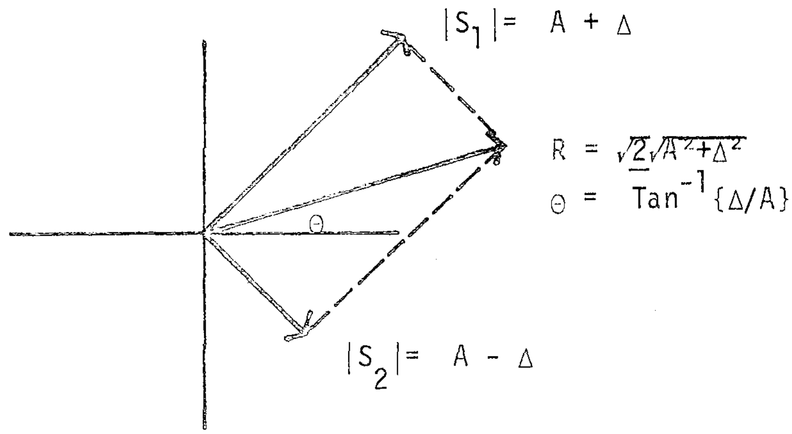
Figure 12. Portable Unit Transmitter Schematic

FOLDOUT FRAME 1

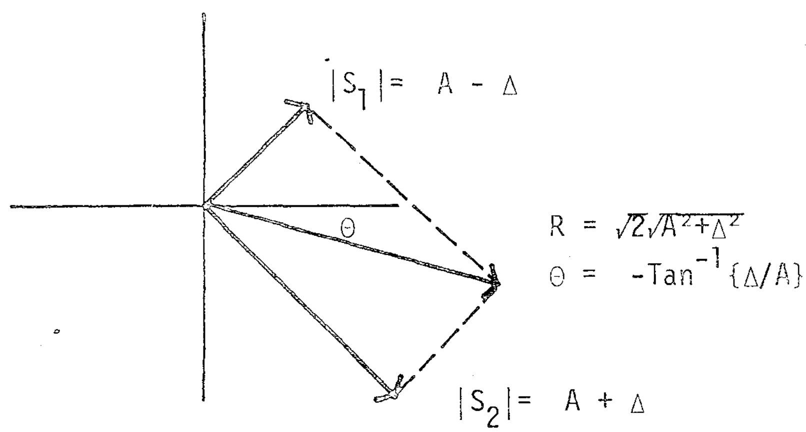
FOLDOUT FRAME 2



No Modulating Signal



Positive Modulating Signal



Negative Modulating Signal

Figure 13. Phasor Representation of Phase Modulation.

implemented using a matched pair of field effect transistors. FETs are used because their high input impedance will not load the phase splitting network. Each FET is amplitude modulated by applying a modulating signal to the source. The signal applied to the source of one FET is made 180 degrees out of phase with the signal modulating the other FET so that push-pull operation can result. The drain currents are added in the tank circuit which is common to both drain terminals (Figure 12).

This implementation has two major advantages over single transistor phase modulators. One is that the available phase deviation is greater; the second and more important advantage is that the resulting phase modulated signal has considerably less amplitude variation. This is important when frequency multipliers are used, since, when frequency multipliers are operated Class C, they are very sensitive to amplitude variations.[18]

Transmitter power switching is achieved by a FET switch shunting the phase modulator output circuit. This implementation was used primarily to avoid delays caused by oscillator starting time. The oscillator is thus in operation continuously. Other advantages of this power switching scheme include:

1. no charging power surges are required from the voltage supply; and
2. no DC power loss occurs across solid state switching elements.

The FET used is a N-channel depletion mode JFET. When a logical 1 (+5 v) is applied to the gate, the channel resistance is low and therefore shorts the tuned circuit. Since the remaining part of the transmitter is operated Class C, no power will flow to these circuits. When a digital 0 (-5 v) is applied to the gate, the channel resistance is high and consequently will allow transmitter operation.

The first and second frequency triplers that follow the phase modulator are very similar. The transistors are biased beyond cutoff so that they operate Class C. The collector load consists of a double tuned capacitively coupled circuit.[19] The double tuned circuits were designed to be slightly under-coupled so that attenuation of undesirable harmonics would be good. The output of the double tuned circuit was connected to the input of the next stage by means of a small capacitor which functioned as both a coupling element and an impedance matching element. A frequency doubler which produces the 80 MHz final RF frequency follows the second frequency tripler; the output of the doubler is fed to a Class C amplifier which drives a quarter wave antenna. The output amplifier is matched to the antenna by a capacitive voltage divider. All individual stages comprising the transmitter have power supply decoupling as shown in Figure 12.

4. PCM Encoder

The front end of the PCM encoder consists of a switch network, calibrate signal generator, and signal conditioners (Figure 14). The switches preceding each signal conditioner are CMOS bilateral switches connected in a double pole double throw (DPDT) configuration. The switches are operated by OPERATE and CALIBRATE digital signals. An OPERATE signal switches the differential inputs of the signal conditioners to electrodes situated on the astronaut's body. A CALIBRATE command connects the calibrated generator to the inputs of the signal conditioners.

The calibrate signal generator is basically a CMOS astable multivibrator set for approximately 5 Hz. A precision 1 mv peak-to-peak differential signal is produced by the resistor divider network and the Zener diode regulation. The extra inverters and the dual resistor divider are used to cancel common mode voltages.

The signal conditioners each have gains of 1000 and are designed to handle up to a ±2.5 mv differential input signal. DC offset voltages are no problem with this circuit because of the AC coupling between the buffer amplifiers and the differential amplifier. The use of large values for the coupling capacitor and resistor in series with the capacitor gives a low frequency cutoff of 0.03 Hz which is adequate for most physiological signals.[9]

Five of the seven instrumentation amplifiers are connected to a multiplexer integrated circuit (Z_2, see Figure 15). The multiplexer consists of eight channel select decoders and eight select switches. The channel select decoder is addressed by a word counter (Z_6). The word counter also addresses additional CMOS switches (1/2 of Z_8) to submultiplex the remaining two physiological channels. The remaining multiplexer inputs are tied to system common corresponding to the zero reference channel and frame synchronization period. The multiplexer output is connected to the pulse width modulator. The pulse width modulator consists of a sweep generator and voltage comparator (Z_4).

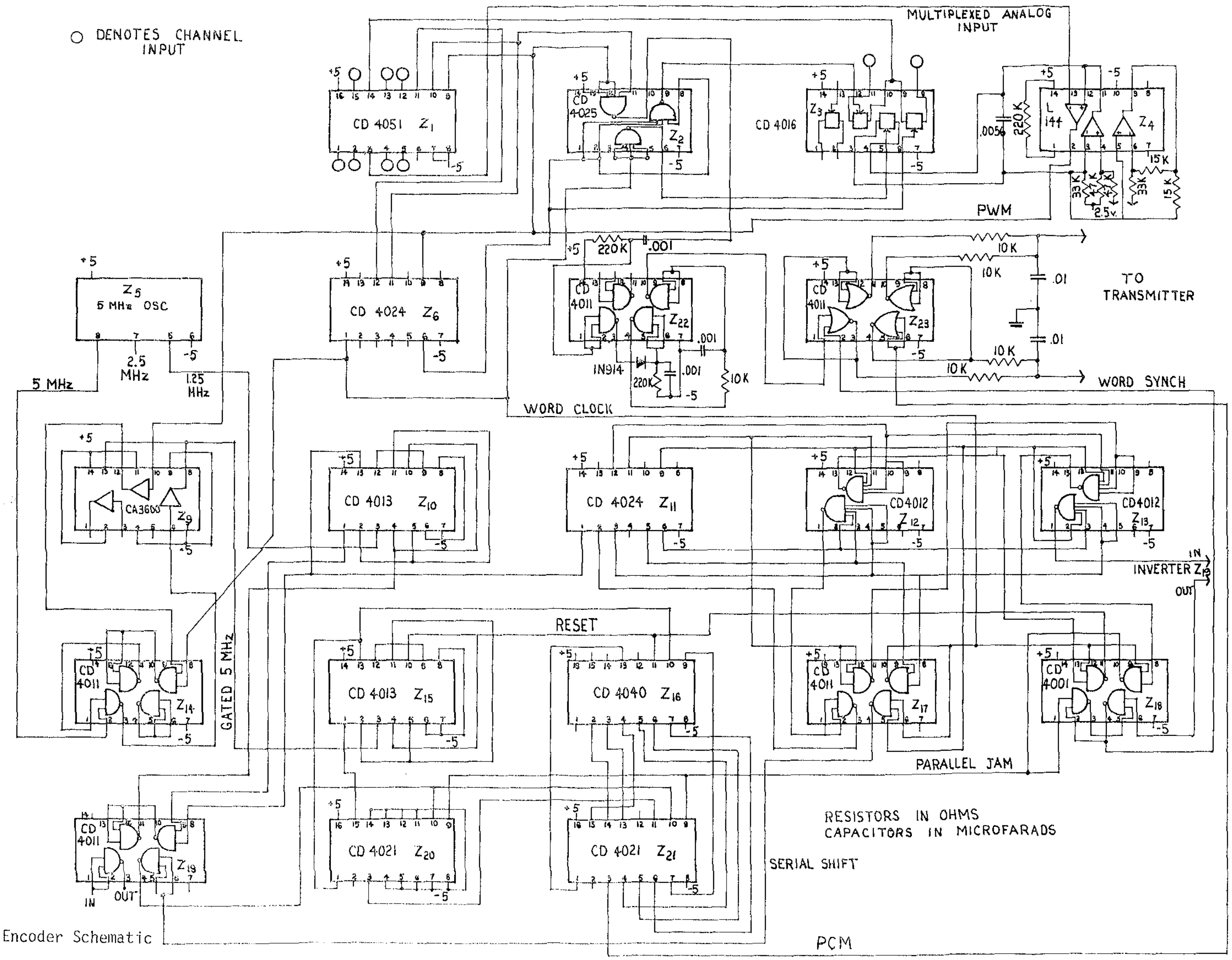


Figure 15. PCM Encoder Schematic

FOLDOUT FRAME /

FOLDOUT FRAME

The sweep circuit is a constant current source which drives an operational amplifier integrator. The sweep is clocked by the word clock signal (Figure 16) by discharging the integrating capacitor C_I. The output of the voltage comparator is high until the sweep voltage equals or exceeds the multiplexer output. The output of the voltage comparator and the word clock are applied to a logical AND gate (1/4 of Z_{14}) resulting in a pulse width modulated (PWM) signal (Figure 16).

The pulse width signal gates (1/4 of Z_{14} and 1/4 of Z_9) a 5 MHz signal to a counter (Z_{15} and Z_{16}). The oscillator is a hybrid CMOS crystal oscillator which has outputs of 5 MHz, 2.5 MHz, and 1.25 MHz. After the gated oscillator pulses are counted, the contents of the counter are parallel jammed into a shift register (Z_{20} and Z_{21}). The counter is then reset before the next gated signal is available. While the next signal is being counted, the contents of the shift register are shifted out serially to the transmitter. A timing diagram of the various control signals during each word period is given in Figure 17. These control signals are generated by the digital integrated circuits Z_{10} through Z_{13}, Z_{17}, and Z_{18}.

The serial data, together with the word synchronization data (Figure 16) are combined to produce two PCM signals which differ by 180 degrees. It should be noted that there is a one word delay between sampled data and the PCM word representing the sampled data. This signal is applied to a low pass filter (cutoff at 2 KHz) and subsequently to the push-pull inputs of the phase modulator on the

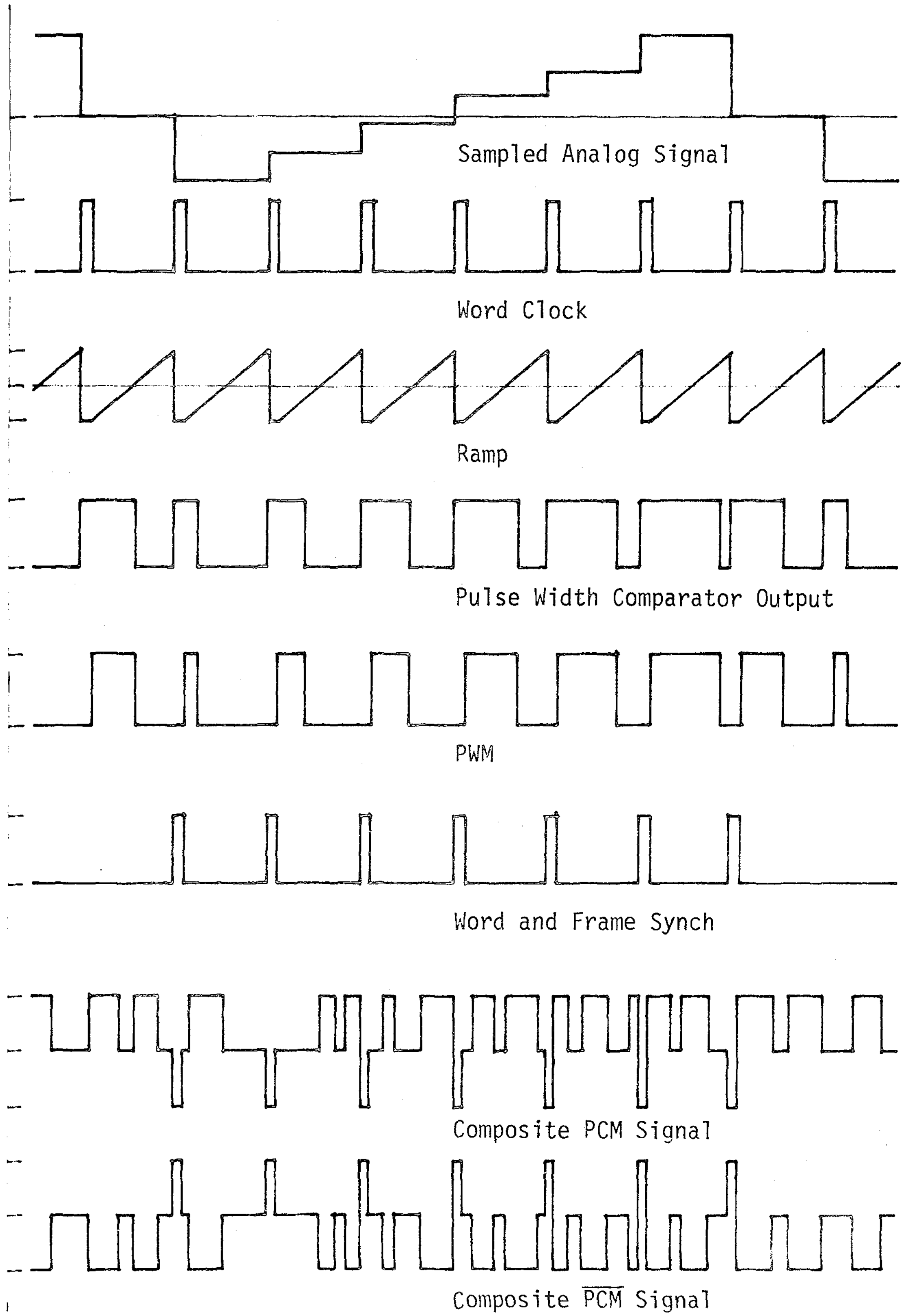


Figure 16. PCM Encoder Timing Diagram

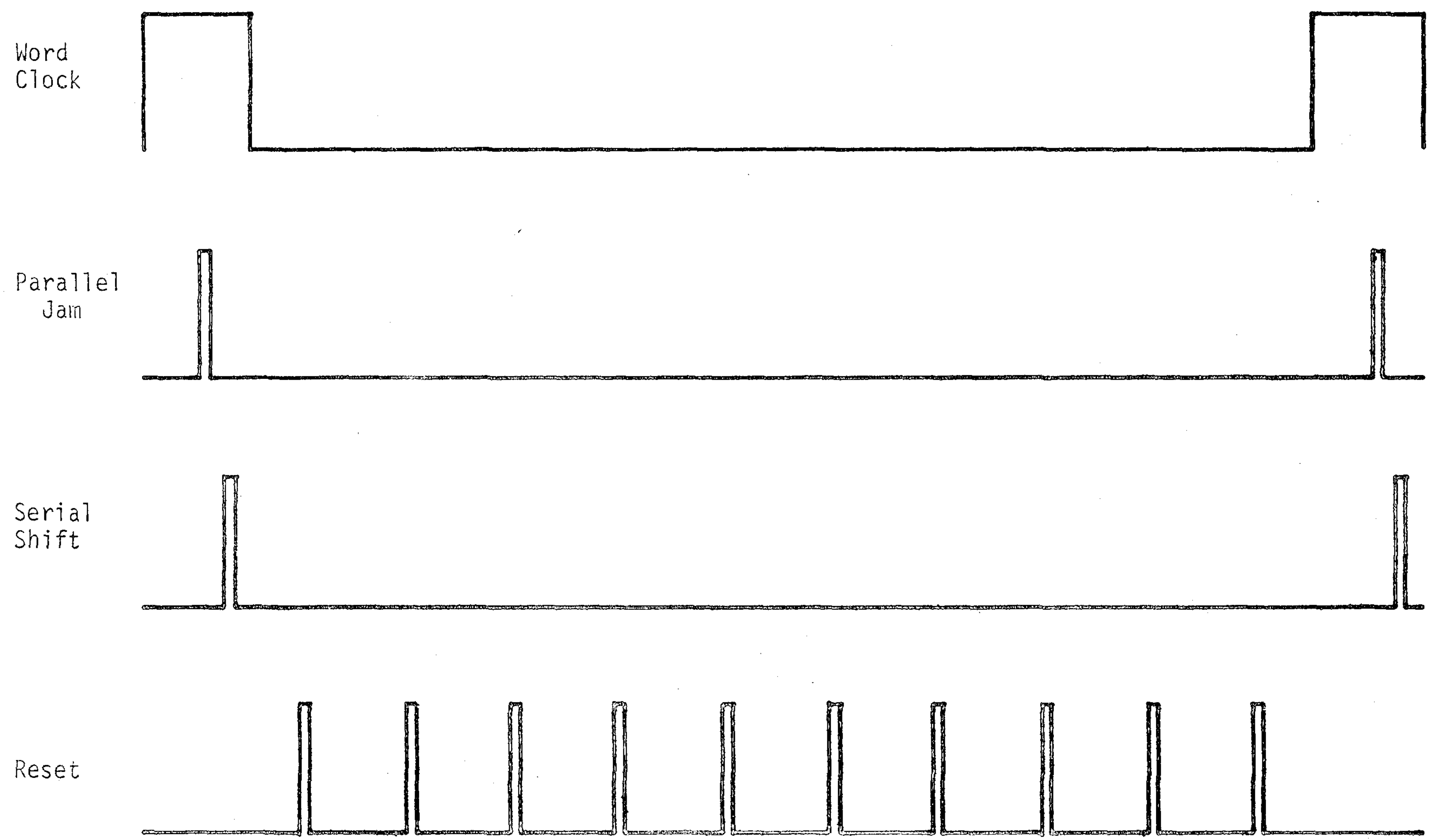


Figure 17. PCM Encoder Timing Diagram During One Word Period

transmitter board.

The PCM encoder shown in Figure 15 is the same as the one designed by NASA Ames Research Center with two exceptions. The CMOS quad NOR gate (Z_{23}) was added to interface with the transmitter. The second modification consisted of changing the synchronization format slightly to be compatible with the decoder when connected by the "pseudo-indirect" FM link. The original synching signal consisted of a one bit duration negative word synch pulse between regular data words and a one word duration negative pulse separating the individual frames. The high pass response inherent with "pseudo-indirect" FM results in considerable distortion in the long duration frame synch signal which causes unsatisfactory operation of the decoder. Consequently, a one shot multivibrator (Z_{22}) was incorporated in the encoder such that the word synch pulse immediately preceding the frame synch slot would be skipped. The skipped word synch pulse has the same effect as the original format, so that no modifications of the decoder were required in this respect.

C. Ground Based Station

The transmitter and receiver are commercial units and will not be discussed here. A general description of the transmitter and receiver is provided in the GBS description in Chapter II.

1. PCM Decoder

The PCM decoder is identical with the system that NASA Ames Research Center designed to be compatible with the PCM encoder, with one exception. The high pass response that resulted with "pseudo-indirect" FM caused degraded performance of the decoder.

The degraded performance was a result of the difficulty involved in determining digital levels. In consequence, additional wave shaping circuits were required at the input of the decoder.

The transfer function of a first order high pass filter is

$$H_{hp}(s) = s/(s + \omega_1) , \tag{13}$$

where ω_1 is the low frequency cutoff.

Ideally, the transfer function describing the received signal as a function of the encoder signal would be

$$H(s) = \frac{E(s)}{R(s)} = 1, \tag{14}$$

where E(s) and R(s) denote the encoder signal and received signal, respectively. In other words, the input signal and output signal would be identical.

The desired transfer function would be obtained if the high pass function (Equation 13) were cascaded to a transfer function which exhibited the reciprocal response. That is, if the received signal was operated on by the reciprocal high pass function,

$$H_{hp}^{-1}(s) = (s + \omega_1)/s = 1 + \omega_1/s, \tag{15}$$

the result would be

$$R(s) = s/(s + \omega_1) \cdot (s + \omega_1)/s \cdot E(s) = E(s). \tag{16}$$

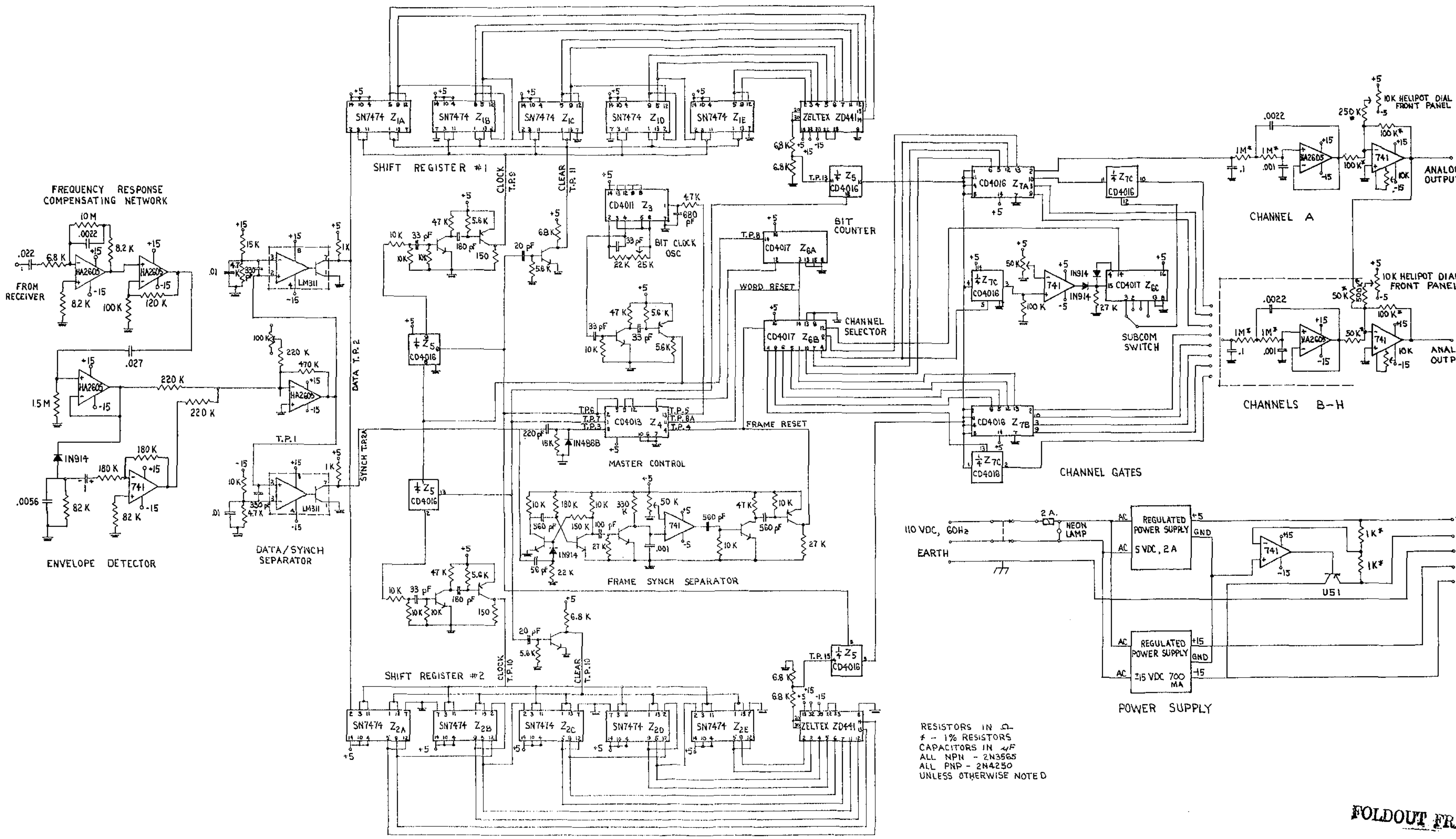
The implementation of the transfer function given in Equation 15 is straightforward. The transfer function can be realized by an integrator and adder stage. The integrating capacitor and input resistor are chosen to provide the ω_1 integrating constant. The actual implementation can be realized using a single operational amplifier configuration.[20]

The response correcting network improved the waveshape of the PCM data considerably but introduced an additional problem. The circuit overemphasized the frame synch period component (590 Hz), resulting in a relatively slow variation in the levels of the digital data. Component values were adjusted in an attempt to reduce this effect in the response correcting network, but no improvement could be obtained. Additional signal processing was required to prevent this variation in level from affecting the following voltage comparators. An envelope detector was used to detect the undesirable variation; the output of the envelope detector was then inverted and added back to the original signal, resulting in cancellation of the undesirable component. This procedure was chosen, rather than a notch filter, because no circuit adjustments were required.

The compensated PCM signal is then fed to two voltage comparators where NRZ data and synchronization data are separated. One word of NRZ data is loaded into a shift register (Z_1) by ten clock pulses produced by a CMOS astable multivibrator (Z_3). The CMOS astable multivibrator is enabled by the preceding word synch pulse and is held on for ten clock pulses by a D flip-flop (1/2 of Z_4) which is in turn reset by a bit counter (Z_{6A}). The reset pulse from the bit counter also updates a channel select counter (Z_{6B}).

The tenth bit indication from the bit counter also clocks a second D flip-flop connected as a toggle flip-flop (1/2 of Z_4). This second flip-flop activates switches so that the next word will be loaded into a second shift register (Z_2). While the second word is being loaded into the second shift register, the first word is held by the first shift register for a one word period and is converted to an analog value by a high speed DAC. The resulting analog value is demultiplexed to the proper output amplifier by switches (Z_{7A} and Z_{7B}) controlled by the channel select counter (Z_{6B}). At the end of the second word period, the first shift register is cleared and is ready to accept the next word. Meanwhile, the second shift register holds the second word which is converted to an analog level and demultiplexed. The timing diagram for the systems is shown in Figure 18. A complete schematic of the PCM decoder is given in Figure 19.

A skipped word synch pulse indicates a frame synch signal. A frame synch separator generates a frame synch pulse essentially by comparing time intervals between word synch pulses. Each word synch pulse discharges capacitor C_S (Figure 19) to ground. Subsequently, the capacitor charges toward a positive five volts through a 330 kohm resistor. This voltage is compared against a positive DC voltage by an operational amplifier used as a voltage comparator. If the R_SC_S time constant is properly set, the capacitor voltage will exceed the DC voltage only in the case of a skipped word synch pulse. When



FOLDOUT FRAME

FOLDOUT FRAME

the capacitor voltage exceeds the DC voltage, the voltage comparator output will switch from a low voltage (-5 volts) to a high voltage (+5 volts). This step voltage signal is then differentiated, producing a short frame synch pulse. The frame synch pulse is used to reset the channel selector counter (Z_{6B}) and master control (Z_4).

The submultiplexed channel is demultiplexed by use of a counter and a reset scheme which closes the demultiplexing switch every other frame count. To ensure that the desired submultiplexed channel is being demultiplexed, the two physiological channels are required to have different peak values. The voltage comparator is set to differentiate between the two peak values and to reset the counter if the wrong channel is being submultiplexed. The fully demultiplexed channels are filtered by an active low pass filter with high frequency cutoff of approximately 100 Hertz.

2. Control Signal Generator and Audio Circuits.

The tone generator, tone reeds, tone select logic and audio stages in GBS are located on a single board. A schematic of the control circuits on this board is shown in Figure 20.

The audio portion of this board consists of a switch network, an audio buffer preamplifier, and an audio output stage (Figure 21). The switch network is used to inhibit the signal to the following audio stages when PCM data are transmitted. The word synch output on the decoder board (digital levels of +5 v and 0 v) is interfaced to the

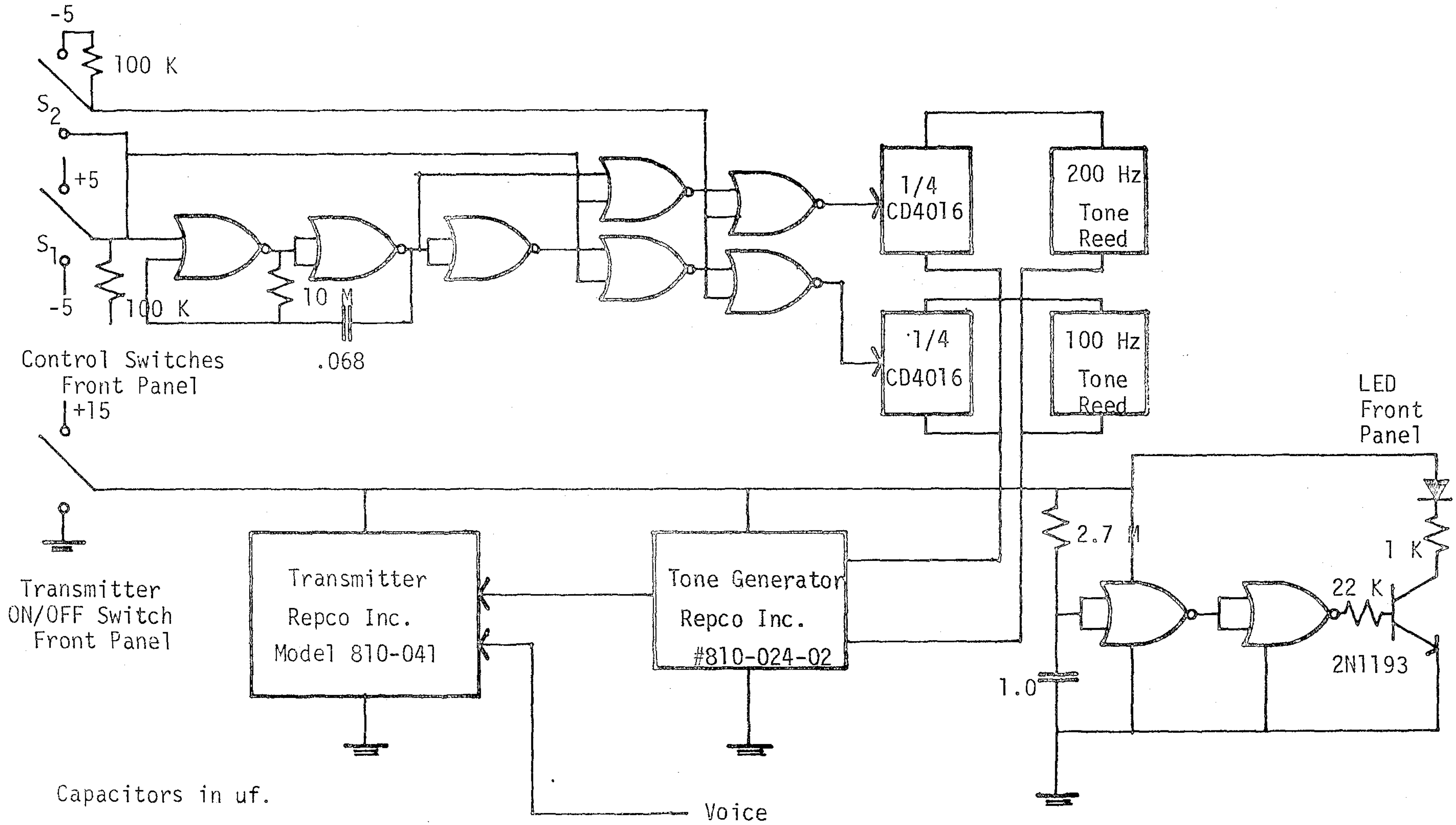
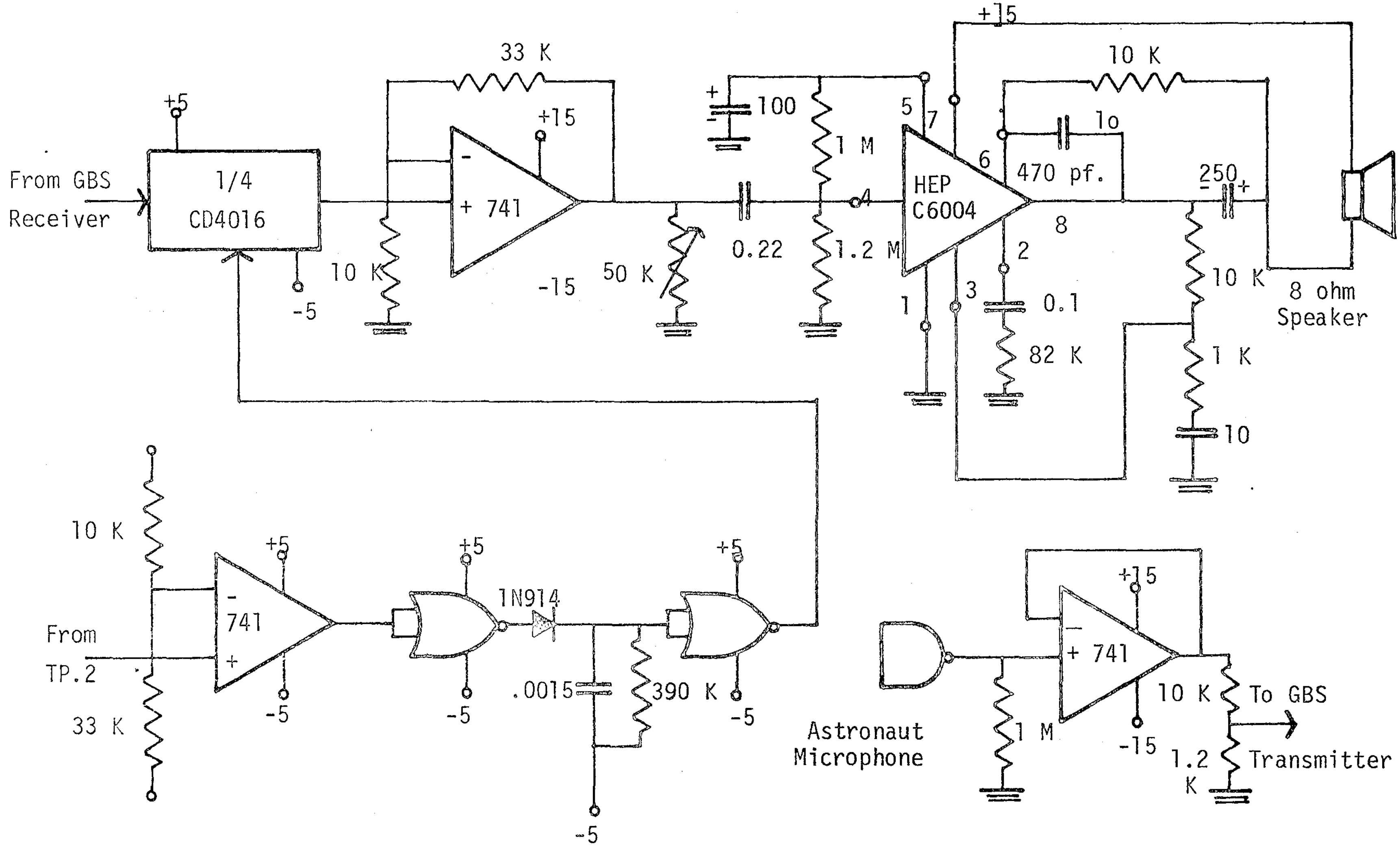


Figure 20. GBS Tone Control Schematic



All Capacitors in uf. unless otherwise noted.

Figure 21. GBS Audio Schematic.

CMOS circuitry (digital voltage levels of -5 v and +5 v) using a voltage comparator, and then fed to a retriggerable one shot multivibrator. The output of the one shot is used to control the audio switch. When the PCM signal is present, the switch is opened so that the undesired signal will not be heard over the speaker. When the PCM signal is terminated, the switch will close and allow voice signals to be heard. The buffer preamplifier is an operational amplifier connected in a voltage controlled voltage source configuration and has a gain of three. The audio output stage is a linear integrated circuit, one watt, audio amplifier driving a 0.5 watt, 8 ohm speaker. The audio amplifier is connected according to the application information supplied with the Motorola HEP C6004 integrated circuit.

The tone control circuitry consists primarily of an astable multivibrator with a frequency of 1 Hz. Various inhibit and enable signals originating from the tone select switches are provided. When switch S_1 is set for a logical 1 (+5 v), it inhibits multivibrator operation and allows one tone to be selected. With S_1 set in the inhibit position, switch S_2 determines which of the two tones will be activated. With switch S_1 set in a logical 0 position (ground potential), the astable multivibrator is allowed to operate. Because of the conditional connection of S_2 to S_1, S_2 has no control in this mode.

A delayed indication of GBS transmitter power is provided on this board. The purpose of this indication is to inform the GBS controller that enough time has elapsed since the activation of tone transmission for the receiver in the portable unit to be locked on. The circuit essentially measures capacitor charging time. When the capacitor voltage reaches the CMOS transmission level, an LED on the GBS front panel is turned on. The delay between initial tone transmission and LED indication is about 3 sec.

A microphone buffer stage is provided on this board. The circuit is used as an impedance converter so that the 1 megohm microphone source is matched to the 5 kohm input impedance of the transmitter. A voltage divider was used to reduce the microphone signal to the proper level required by the transmitter (7 mv RMS for full modulation).

CHAPTER IV

DISCUSSION OF RESULTS

The GBS unit and the portable unit are shown in Figures 22 and 23, respectively. The system operates within the specifications established by NASA JSC. For the complexity of the system, the power requirement of the portable unit is quite low. A number of techniques for improving system performance became apparent in the last stages of design and subsequent testing; these modifications are presented as suggestions in this chapter.

A. Results

The system was designed to have the capability for seven channels of physiological data. However, only three channels of ECG data were used in testing. Other physiological data, such as temperature and respiration, were not used because of the requirement of additional signal conditioning. Such signal conditioning can be incorporated into the system, but was not required in the specifications.

An illustration of a typical signal generated by the PCM encoder is shown in Figure 24(a). The PCM signal received by the GBS receiver is shown in Figure 24(b). The high pass response indicated by the received signal is a result of the "pseudo-indirect" FM format. The output of the response correcting network is shown in Figure 24(c). A noticeable improvement in noise margin is exhibited by this signal.

Strip charts of three channels of ECG data are shown in Figure 25(a). These are compared to the signal at the output of the signal

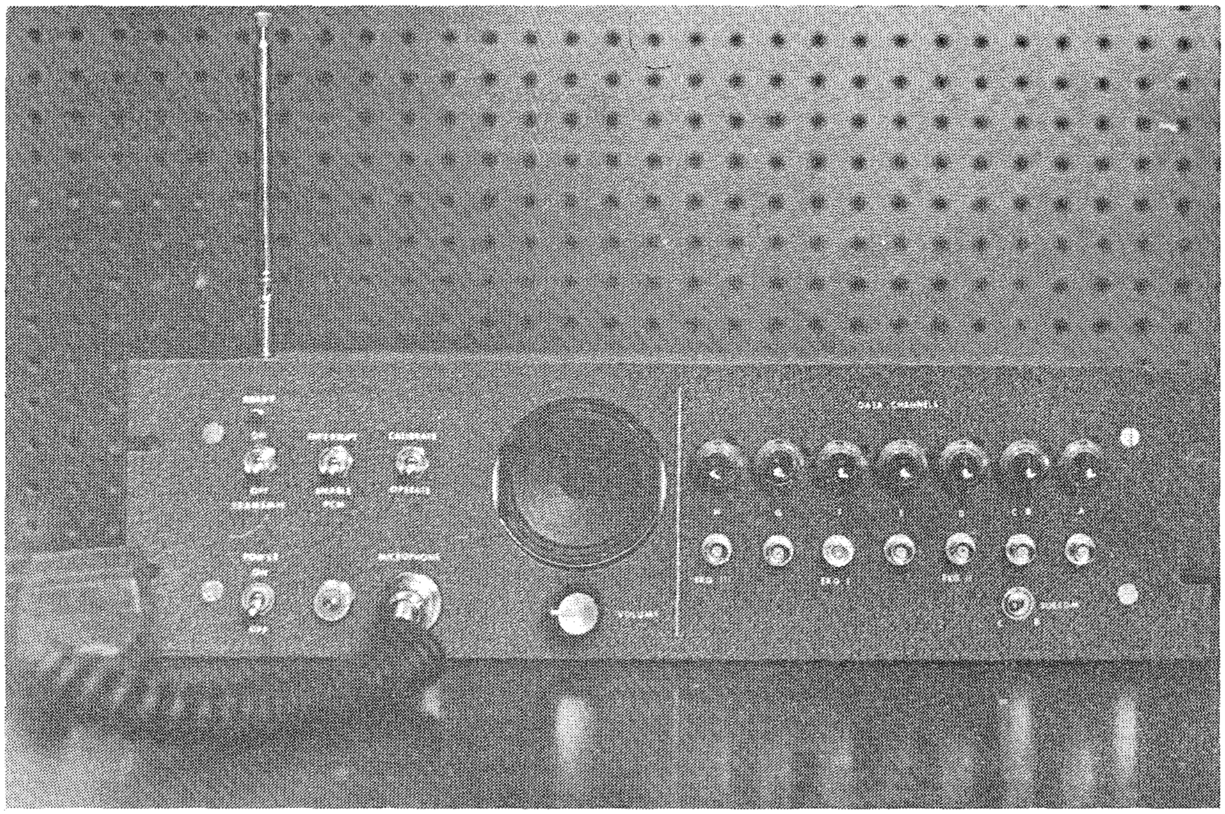


Figure 22. The GBS Unit

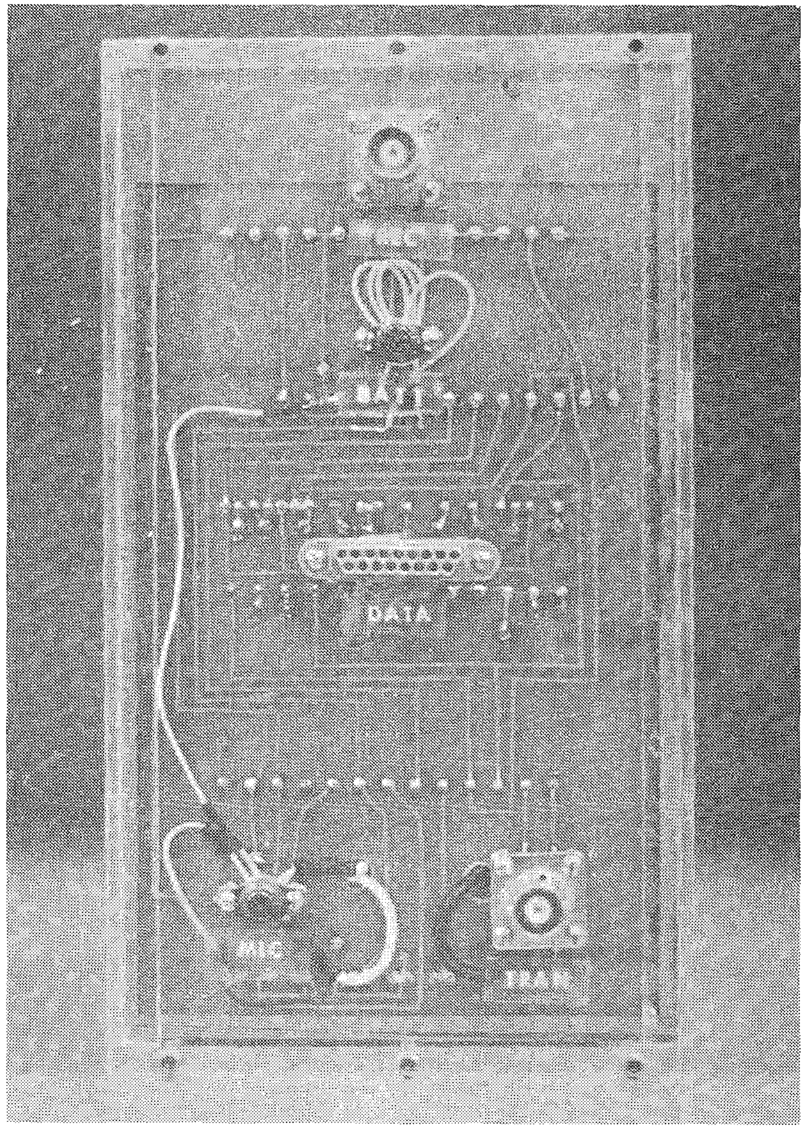


Figure 23a. Front View of the Portable Unit

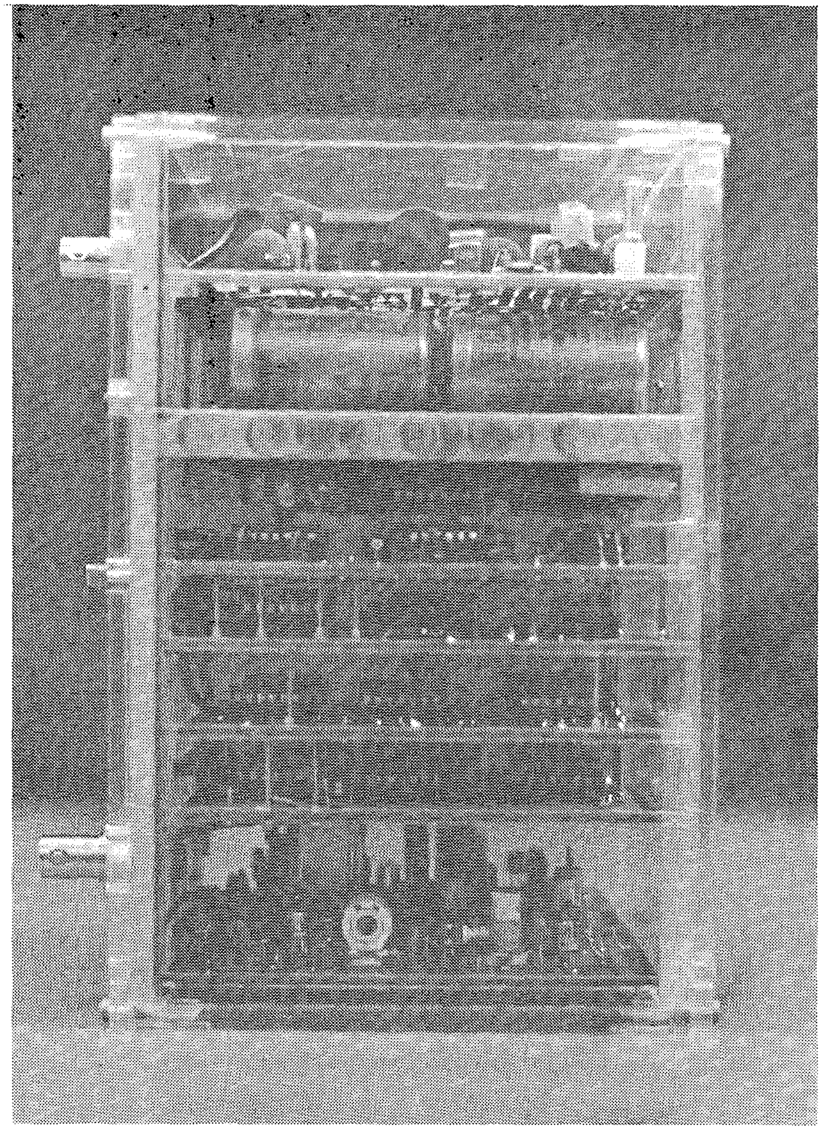


Figure 23b. Side View of the Portable Unit

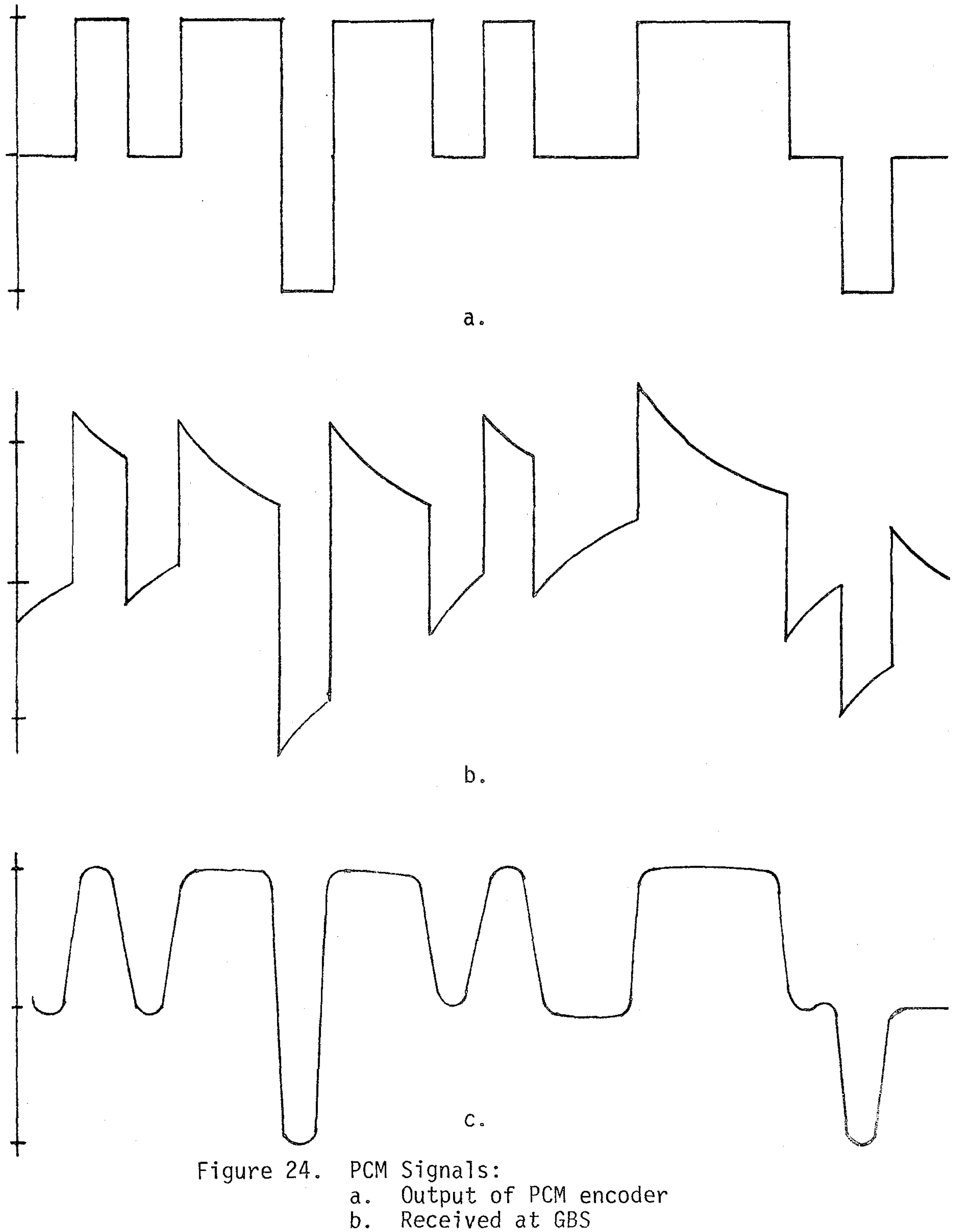
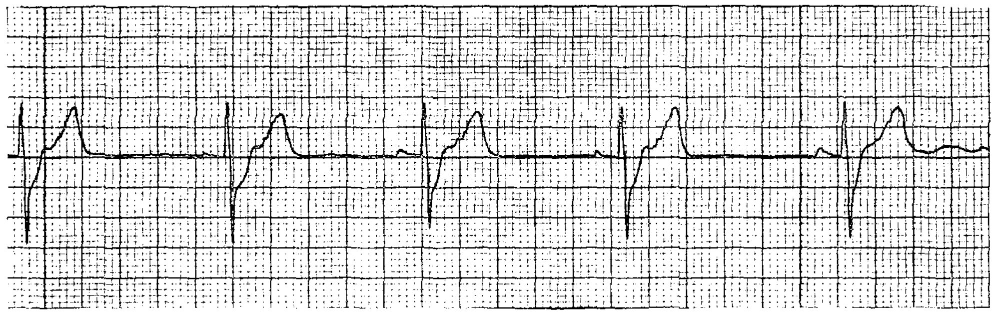
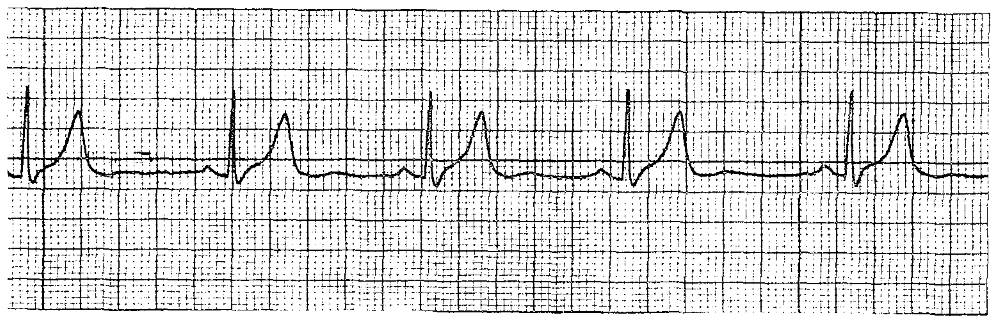


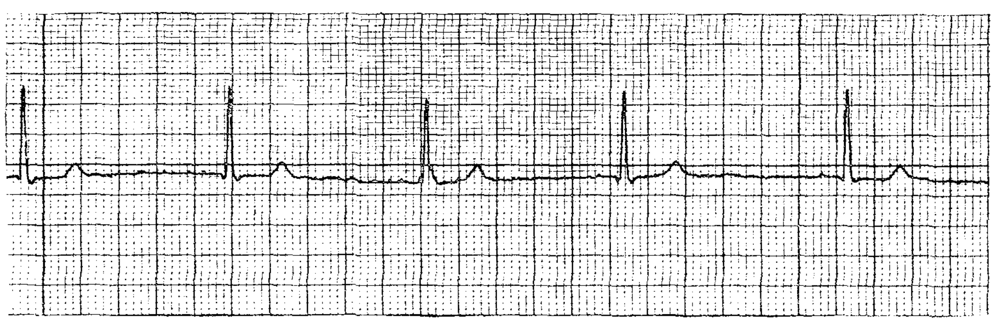
Figure 24. PCM Signals:
a. Output of PCM encoder
b. Received at GBS
c. Delivered by the compensating network



Lead I



Lead II



Lead III

Figure 25a. Decoded Telemetry ECG Data
Horizontal Scale: 25 mm per sec;
Vertical Scale: 200 mv per mm

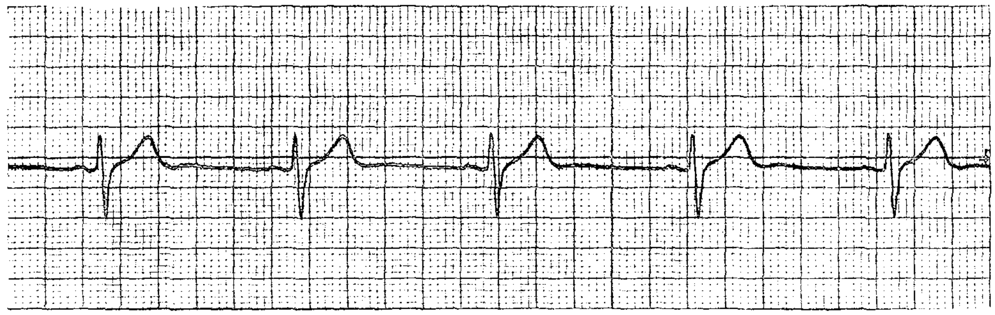
conditioners in the portable unit (Figure 25(b)). High quality reproduction of the heart wave is noted.

The ranges of the system's RF links are adequate. Reliable transmission of both PCM signal and voice is achieved at a distance of 100 ft from GBS. Further range can be demonstrated; however the received signal becomes somewhat noisy at greater distances.

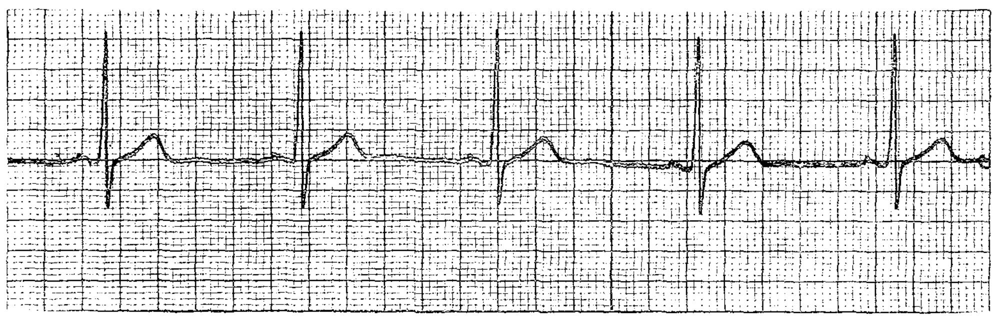
When all subsystems of the portable unit are activated, a total current drain of 65 ma is required of the battery source. The batteries used are rated at a capacity of 1.2 ampere-hours. This capacity corresponds to 20 hours of continuous operation before a battern recharge is required. In actual use, however, the battery life will be greater since subsystems will not be required continuously.

B. Proposed Modifications

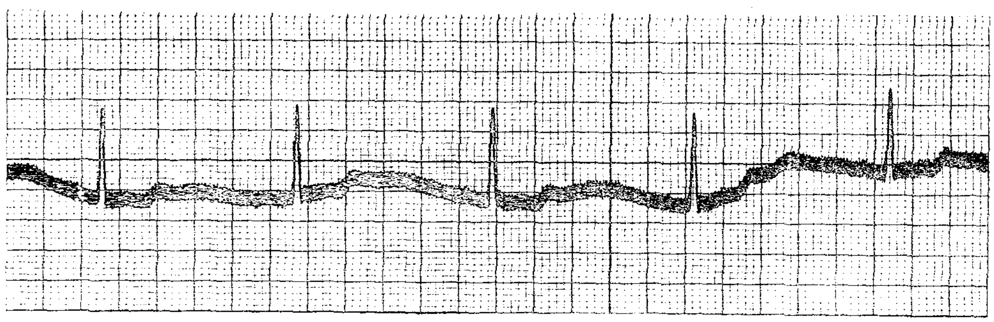
The problems encountered with the portable unit's transmitter are basic ones. The combined specifications of crystal control and wide deviations required by the PCM data signal are mutually exclusive. The use of a "pseudo-indirect" FM system increased deviation but caused loss of valuable low frequency information. Transients caused by the high pass filter response are very random in nature, that is, the voltage spikes and voltage droop discharge times vary with the changing position and duration of the NRZ data (Figure 25). The resulting effect is a decrease in the noise margin of the PCM digital signal. For a digital signal, the noise margin is defined as the difference in the lowest voltage representing a logical 1 and the highest voltage representing a logical 0. To circumvent



Lead I



Lead II



Lead III

Figure 25b. Transmitted Telemetry ECG Data
Horizontal Scale: 25 mm per sec;
Vertical Scale: 100 mv per mm

this problem, additional waveshaping circuits were used (see Chapter III).

Ultimately, the best approach to solve the wide bandwidth problem would be a redesign of the transmitter and possibly a change in the modulation. Simultaneous transmission of astronaut's voice and PCM data should be included in such a redesign.

The problems associated with simultaneous transmission of voice and PCM data have been discussed previously. The resulting frequency spectrum of the RF signal must consist of two separated RF spectra. Implementation of such a scheme requires essentially two separate transmitters which possibly would be matched to a single antenna by use of a directional coupler. Two transmitters cause increased power consumption and size; consequently, a re-evaluation of the design approach and of the operating specifications of the transmitter must be made.

The redesign of the transmitter includes

1. low power consumption;
2. small size;
3. reduced number of RF coils for hybridization purposes; and
4. a solution of the bandwidth problems associated with PCM data.

To keep size and power consumption down, at least one of the two separate transmitters must be simplified. This simplification must be implemented while causing no degradation in the modulation characteristics.

One approach to the redesign would be to use a different modulation format. One very efficient modulation scheme would be pulse modulation

(PM). In this format, the RF carrier is switched on and off by the modulating digital signal. The exact implementation of a PM transmitter for this application would require additional research. Transmitter power switching or oscillator switching would exhibit inadequate response to the high bit rate PCM data. A better approach would be to use an FET switch shunting a tank circuit of some stage following the oscillator. An alternate approach would be to modulate the emitter voltage of a common emitter class C stage. This would in effect allow class C operation during a low digital modulation signal and inhibit operation at the high digital modulation signal. A PM transmitter of good output power and efficiency could be realized using only three stages: a first stage consisting of crystal controlled oscillator operated at 60 - 70 MHz, a second stage consisting of a frequency doubler with PM modulation, and a final output stage operating at 120 - 140 MHz. Modification of the present PCM tri-state logic to two stage logic would be required for PM transmission.

A two state PCM signal would require a code sequence in place of the present negative synch pulses. To maintain present bit rates, word synch information must be eliminated. This would require the PCM decoder to be able to remain in synchronization with the PCM data for one entire frame period. While this reduction of synchronization information increases the probability of the decoder falling out of synchronization, reliable operation can be realized by using a somewhat more sophisticated decoder synchronization scheme. The frequency of the bit clock which synchronizes with the incoming PCM data can be matched very closely to the crystal controlled PCM bit rate by use of crystal control in the decoder. A frequency tolerance of .01 percent is easily achieved with crystal oscillators and this precision is more than adequate to maintain synchronization for one frame period. Any small difference between PCM bit rate and the bit clock frequency would be corrected by turning the bit clock off for a few cycles during the frame synch word period and restarting it at the proper time to synchronize with the next frame period.

The code used in the frame synch word period would have to be such that it could not be mistaken for a regular data word. In the present PCM encoder implementation, the maximum binary number representing a data word is 1111000000 (960 in decimal). By using a sequence of ten consecutive 1s as the synchronization code, the decoder could recognize this sequence and subsequently induce synchronous operation. A recognition of this code by a sequence detector could be used to turn off the bit clock. Following this code, a less elaborate code would be required to restart the bit clock for the next frame. This bit clock enable code is conditional on the frame code sequence immediately preceding it, and therefore can be any convenient combination of 1s and 0s. It should be noted that 0s separate each word in the frame, thus preventing the possibility of a false detection of a frame synch code.

The sequence detector would be implemented using four flip-flops and appropriate combinational logic. A design procedure for this implementation can be found in the literature describing digital circuits.[21]

The second transmitter used for voice transmission could be either an AM or an FM transmitter. An AM transmitter would allow a simpler implementation; however, more knowledge of the space environment is needed to determine if the more noise sensitive AM RF signal would give satisfactory results. Assuming an FM transmitter is used for astronaut voice communication, a simpler implementation than the present transmitter must be made. A direct FM scheme incorporating a phase locked loop frequency multiplier would reduce size and the number of RF coils; however, no appreciable amount of power would be saved. A regular Class C multiplier stage would be required following the output of the PLL multiplier, since the upper operating frequency of most PLLs are around 30 MHz.[13]

An additional improvement that could be made is a modification of the control subsystem. At present, continuous transmission of control tones, and consequently, continuous operation of the receiver in the portable unit, is required for GBS control. If, instead, a control code were generated such that the receiver would lock for only the time required to receive and properly decode the control information, receiver power consumption would be reduced. In this mode, the receiver would have to sample occasionally to check instruction status.

The modifications listed above would give increased performance over the present system. Power requirements probably would increase due to the extra transmitter required in the portable unit; however, the current drain should not be greater than 100 ma. The physical size of the total portable unit can be reduced considerably by hybridization. It must be noted, however, that portions of the RF circuits are not simply hybridized because of the required RF coils and the higher frequencies involved; possibly gyration or high frequency PLLs could be used for this purpose.

CHAPTER V

CONCLUSION

A seven channel physiological telemetry system was designed, constructed, and tested. Three ECG leads were monitored simultaneously, and two-way voice communication was achieved. The portable unit operated properly under GBS transponder control.

The major design problem was associated with the constraints of a wide bandwidth FM transmitter with crystal control. The tri-state logic format used by the NASA Ames Research Center also compounded this difficulty. However, by using an "indirect FM" scheme and incorporating additional waveshaping circuits at GBS, proper operation of the system was obtained.

In the present form, the telemetry system has capability for seven channels of low level (2.5 mv peak maximum) physiological signals with bandwidth between 0.03 Hz and 100 Hz. Additional signal conditioning and modification of the instrumentation amplifiers gains are required for other signals.

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APPENDIX A

OPERATING PROCEDURE

Once the system has been properly adjusted, the operating procedure is straightforward. Except for the astronaut interrupt control, which consists of a spring loaded switch, the portable unit is automatic and requires no control for proper operation.

The DATA input, MIC (microphone) input, and BATT (battery) input are all keyed connectors. To operate the system, a shorting plug is connected to the BATT input. This plug connects the batteries to the electronic circuits. In order to recharge the batteries, the plug must be removed and a two-lead recharge cable connected, permitting external access. The batteries should be charged at a constant 100 ma rate for 16 hours for full charge.

The operating procedures for the GBS unit are more involved, but also straightforward. Initially, all offset controls should be set at a mid-point (500 on the dial). The SUBCOM channel selector switch determines which submultiplexed channel is decommutated and filtered.

The ON/OFF switch, located in the lower left-hand corner of the panel, switches power to the GBS unit. A neon lamp, to the right of the ON/OFF switch, indicates power status. The microphone input jack is directly to the right of the neon lamp. The microphone itself has a push-to-talk switch; however, voice transmission will not take place unless the transmitter power switch is in the ON position. The transmitter power switch is in the upper left hand corner of the panel, directly above the ON/OFF power switch. Activation of this switch causes an RF signal, modulated by one or both control signals, to be transmitted. Three seconds after transmitter power has been switched on, a red LED indicator, located above the transmitter power switch, turns on and indicates enough time has elapsed for the portable unit receiver to be locked up. Two switches (S_1 and S_2), to the right of the transmitter switch, select which control instruction will be transmitted. Switch S_1, directly to the right of the transmitter power switch, inhibits and enables PCM operation. In the PCM inhibit position (switch S_1 in the up position), switch S_2 (directly to the right of S_1) has no control. When switch S_1 is in the PCM enable position, switch S_2 selects either a CALIBRATE or OPERATE control instruction. In any combination of control switch positions, GBS voice transmission is obtained.

The speaker volume control is located below the speaker. When the PCM ENABLE switch is in the ENABLE position, no signal will be delivered to the speaker.

For proper operation, the GBS receiver signal level should be set to give at least a 2.5 volt peak-to-peak PCM signal.

APPENDIX B

MAINTENANCE

For proper operation of the system, a number of adjustments must be correctly made. These adjustments include transmitter and receiver alignment, tone decoder frequency adjustments, and PCM decoder adjustments.

The portable unit transmitter is aligned by successively tuning each stage, beginning with the phase modulator tank circuit. All stages except the final tripler stage are tuned by varying the slug position in the appropriate coils. A locking insert is provided to maintain slug position. The circuits are tuned to provide a maximum DC voltage at the transistor emitter connection in the final stage. The output tuned circuit can be tuned by using an RF power meter or spectrum analyzer, to provide maximum signal at 80 MHz.

The receiver alignment procedure consists of successively tuning each stage beginning with the RF amplifier. With the GBS transmitter on, each stage is tuned to give maximum 455 KHz signal at the phase-locked-loop detector input. The phase-locked-loop detector has center frequency adjustment. A 2 kohm potentiometer is provided for this purpose. While monitoring the PLL detector output (pin 8), the potentiometer should be varied until the DC level is zero volts with respect to system common.

The tone decoder center frequencies are adjusted by means of 20 kohm potentiometers. These potentiometers are conveniently located facing the rear panel of the portable unit. The 200 Hz center frequency control is located to the extreme right. A 50 kohm potentiometer, directly to the left of the 200 Hz frequency adjustment potentiometer, is used for adjusting the amount of 200 Hz signal fed back to the audio stages for 200 Hz notch operation. With a 200 Hz tone transmitted by GBS transmitter, the 200 Hz bandpass output should be adjusted for maximum signal. The notch adjustment should be made for minimum 200 Hz signal at the audio output stage, which drives the headset. Similarly, the 100 Hz tone decoder is adjusted for maximum 100 Hz bandpass output with a 100 Hz tone transmitted by GBS. The 100 Hz adjustment potentiometer is located to the extreme left on the same board as the 200 Hz adjustment potentiometer.

The adjustments on the decoder board in the GBS unit consist of a DC offset control on the input amplifier and a bit frequency control potentiometer. The DC offset potentiometer is accessible from the rear panel of the GBS unit. The wave form monitored at TP.1 (Figure 18) should be such that the 0 digital level (tri-state logic format) is at ground potential. The bit frequency which synchronizes with incoming data is adjusted by means of a 20 kohm potentiometer. Proper operation is obtained if the frequency is set to within 10 percent of the incoming data rate. The adjustment is made by monitoring the signal at test point TP.8. For proper alignment, the bit pulses should be 19.2 μsec apart.

APPENDIX C

PORTABLE UNIT INPUT CONNECTORS

The battery connector (BATT) and microphone/headset interrupt connector (MIC) are Cannon Type ME 95185-2 connectors. The pin connecting configuration is described in the following tables.

BATTERY INPUT CONNECTOR

Pin Number	Circuit Connection
1	Positive Battery Supply
2	Positive Circuit Connection
3	N. C.
4	Negative Circuit Connection
5	N. C.
6	Negative Battery Supply
7	N. C.

MICROPHONE INPUT CONNECTOR

Pin Number	Circuit Connection
1	Ground
2	Microphone
3	V+
4	Interrupt
5	Headset
6	Ground
7	N. C.

The DATA input connector is an Amphenol Tiny Tim 223-6 connector. The circuit connections are shown in the table below.

DATA INPUT CONNECTOR

Pin Number	Circuit Connection
1	Channel D Differential Input
2	Channel D Differential Input
3	Ground
4	Channel H Differential Input
5	Channel H Differential Input
6	Channel G Differential Input
7	Channel G Differential Input
8	Channel F Differential Input
9	Channel F Differential Input
10	Ground
11	Channel B Differential Input
12	Channel B Differential Input
13	Channel C Differential Input
14	Channel C Differential Input
15	Channel E Differential Input
16	Channel E Differential Input
17	Ground